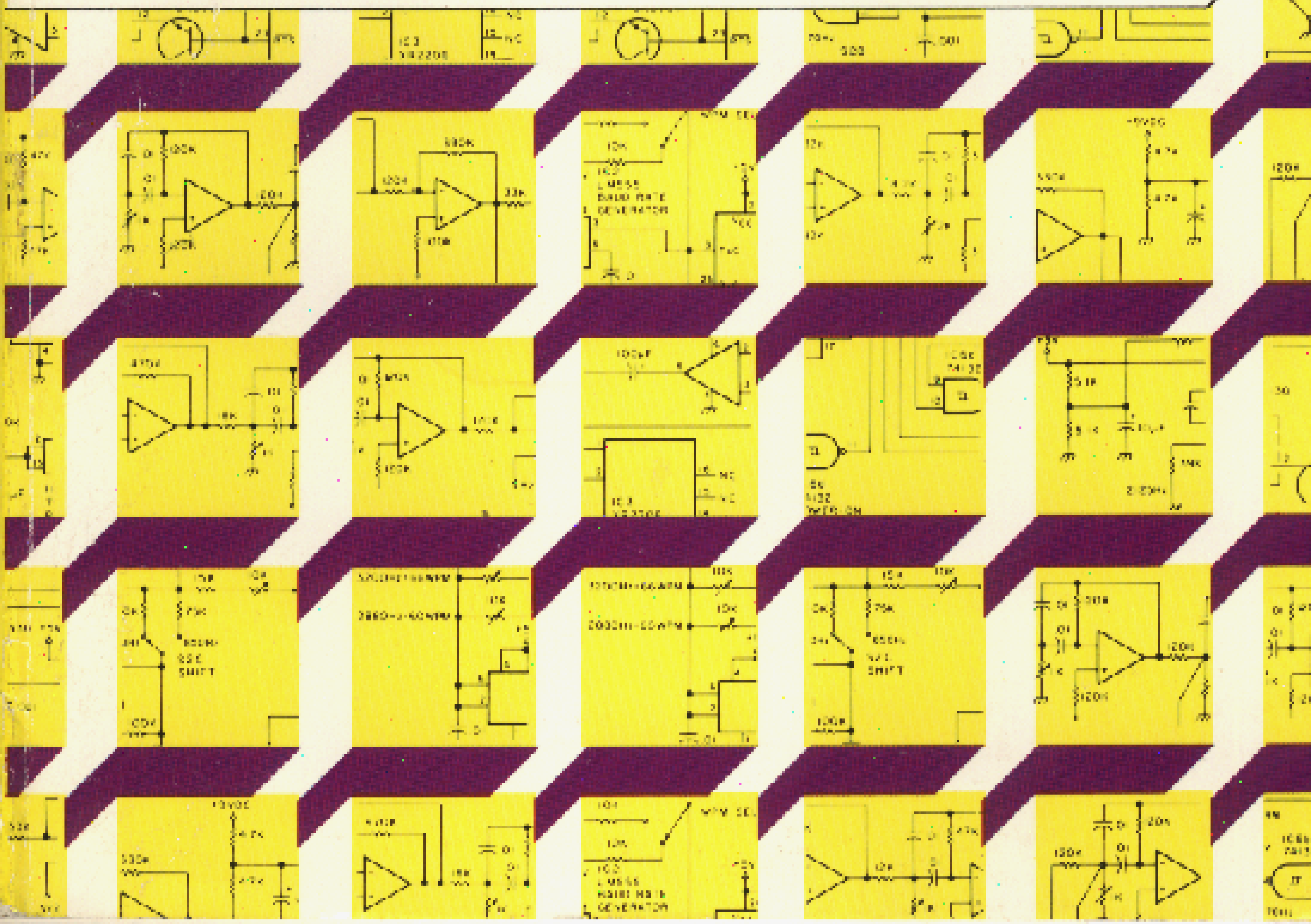
1938

ENCYCLOPEDIA OF ELECTRONIC CIRCUITS

VOLUME 1

RUDOLF F. GRAF



Contents

Introduction

This volume of timely and practical circuits highlights the creative work of many people. Featured here are many circuits that appeared only briefly in some of our finer periodicals or limited-circulation publications. Also included are other useful and unique circuits from more readily available sources.

The source for each circuit is given in the sources section at the back of the book. The bold figure number that appears inside the box of each circuit is the key to the source. For example, the High Stability Voltage Reference circuit shown below is Fig. 93-10. If you turn to the Sources section and look for Fig. 93-10 you will find that Precision Monolithics supplied this circuit from p. 6-142 of their Full Line Catalog.

HIGH STABILITY VOLTAGE REFERENCE

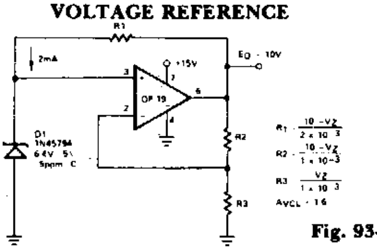


Fig. 93-10

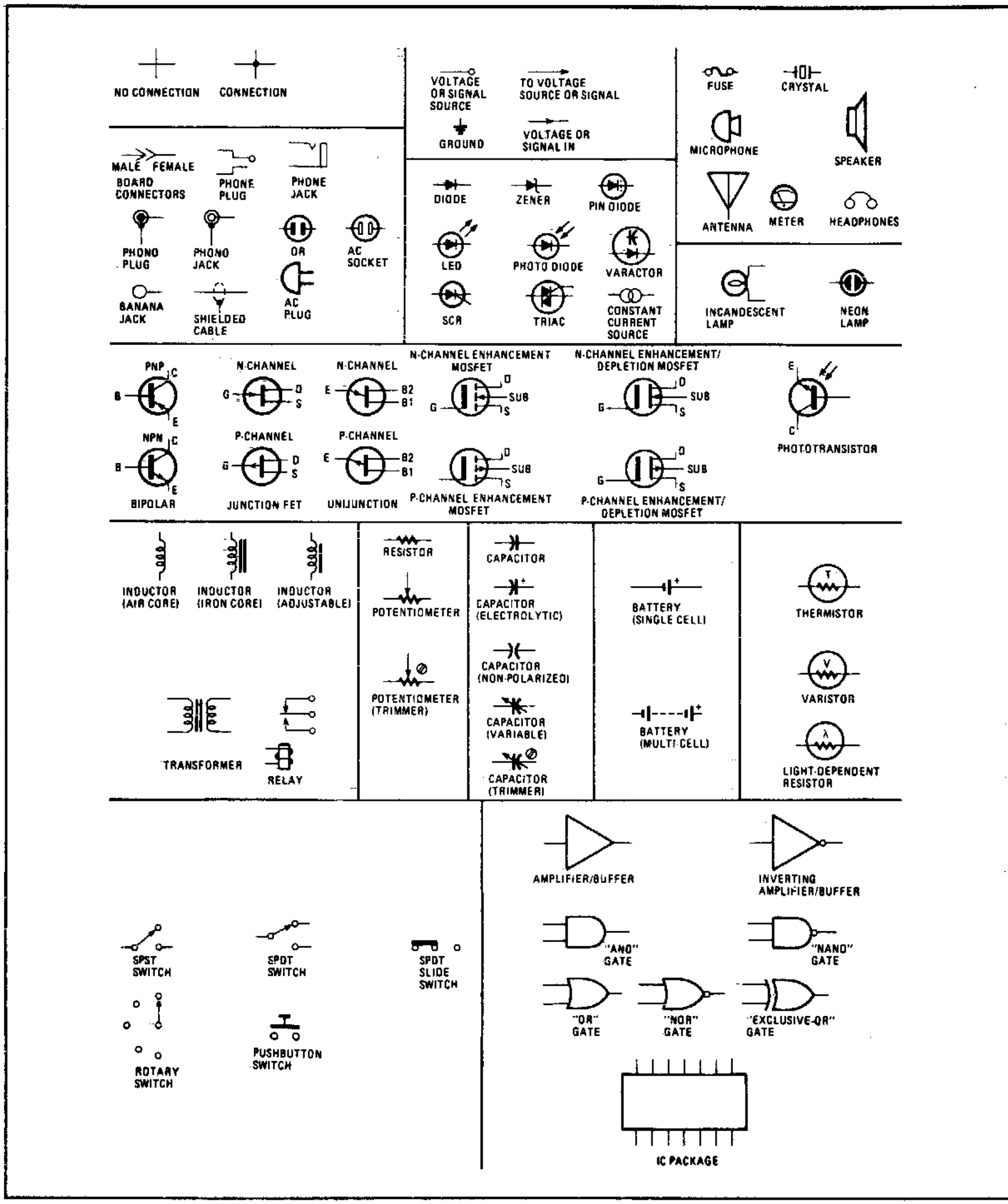
Fig. 93-9: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-25.
Fig. 93-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-142.
Fig. 93-11: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 10-18.

Many circuits are accompanied by a brief explanatory text. Those that do not have text can be readily understood from similar circuits in that chapter, or else they may be too complex to be explained briefly. The sparseness of text is deliberate so as to allow for more circuits which, after all, is what this book is all about.

The Index and Contents will be a time saver for the reader who knows exactly what he is looking for. The first page of each chapter lists the circuits in the order that they appear. The browser will surely discover many ideas and circuits that may well turn out to be most rewarding and great fun to put together.

The Common Schematic Symbols chart will help you identify circuit components.

Common Schematic Symbols



1

Alarms

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

COMPUTALARM

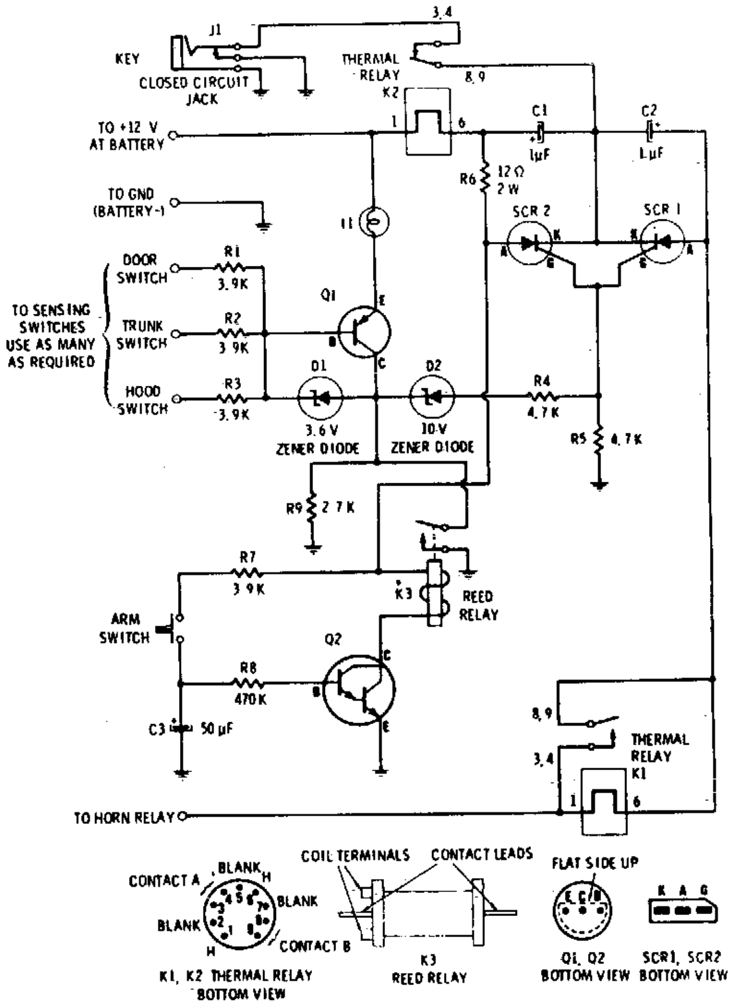


Fig. 1-1

Circuit Notes

The circuit has a built-in, self-arming feature. The driver turns off the ignition, presses the arm button on the Computalarm, and leaves the car. Within 20 seconds, the alarm arms itself—all automatically! The circuit will then detect the opening of any monitored door, the trunk lid, or the hood on the car. Once activated, the circuit remains dormant for 10 seconds. When the 10-second time delay has run out, the circuit will close the car's horn relay and sound the horn in periodic blasts (approximately 1 to 2 seconds apart) for a period of one minute. Then the Computalarm automatically shuts itself off (to save your battery) and re-arms. If a door, the trunk lid, or the hood remains ajar, the alarm circuit retriggers and another period of horn blasts occurs. The Computalarm has a "key" switch by which the driver can disarm the alarm circuit within a 10-second period after he enters the door. The key switch consists of a closed circuit jack, J1, and a mating miniature plug.

AUTOMOTIVE BURGLAR ALARM

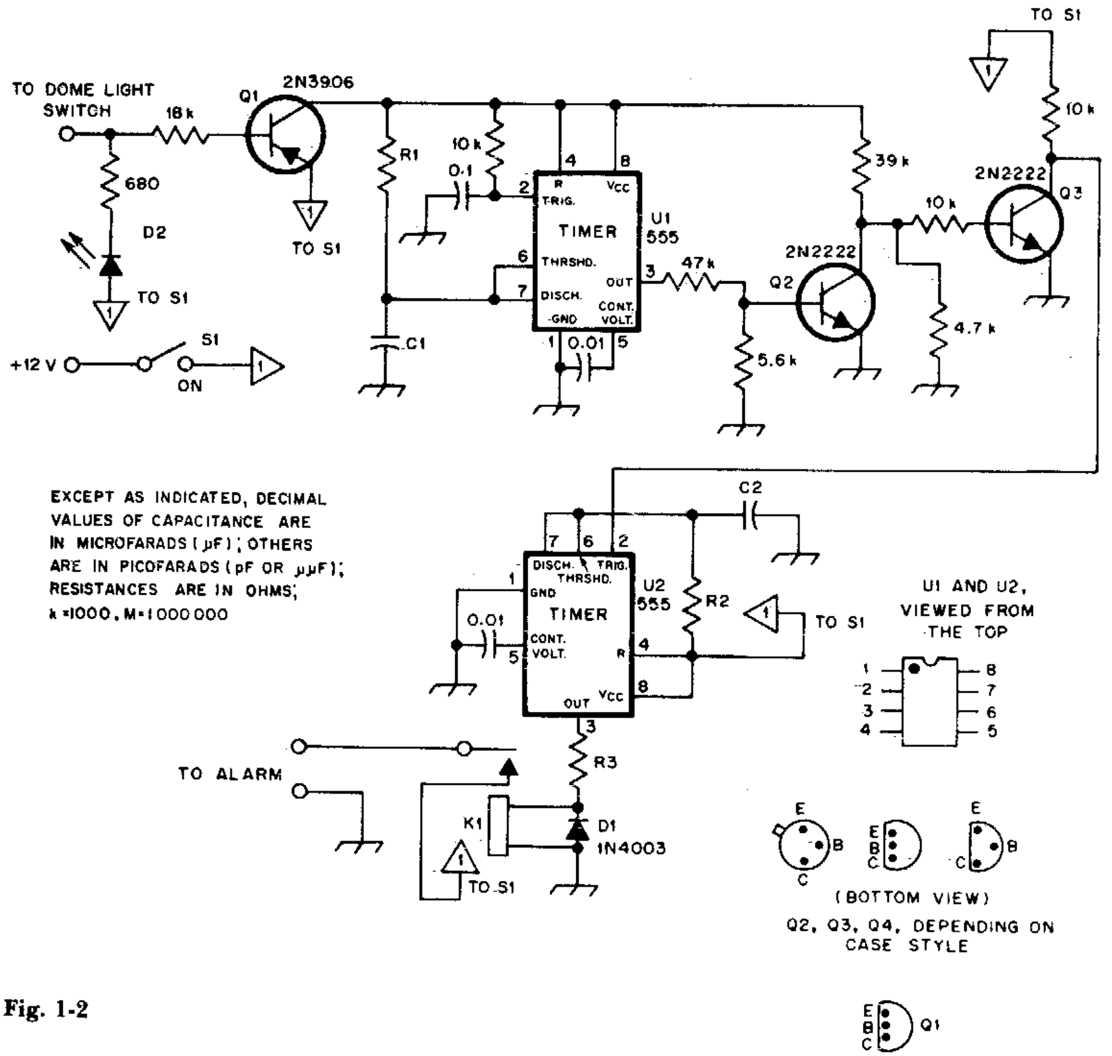


Fig. 1-2

Circuit Notes

Alarm triggers on after a 13 second delay and stays on for 1-1½ minutes. Then it resets automatically. It can also be turned off and reset by opening and reclosing S1.

SECURITY ALARM

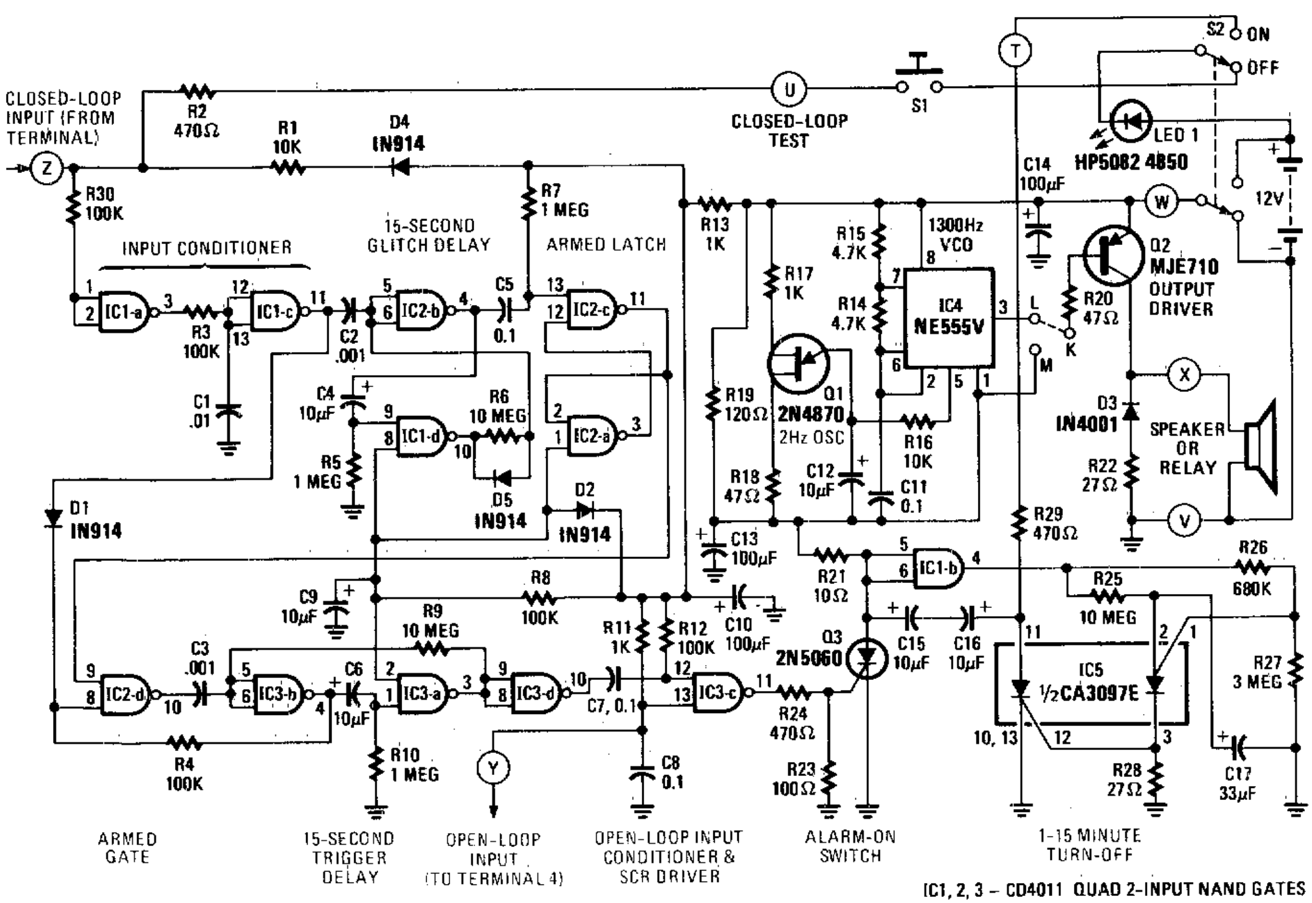


Fig. 1-3

Circuit Notes

This alarm features open- and closed-loop detector and automatic alarm shutoff. Offers 15 second exit/entrance delay. Alarm on time can be adjusted from 1 to 15 minutes.

VEHICLE SECURITY SYSTEM

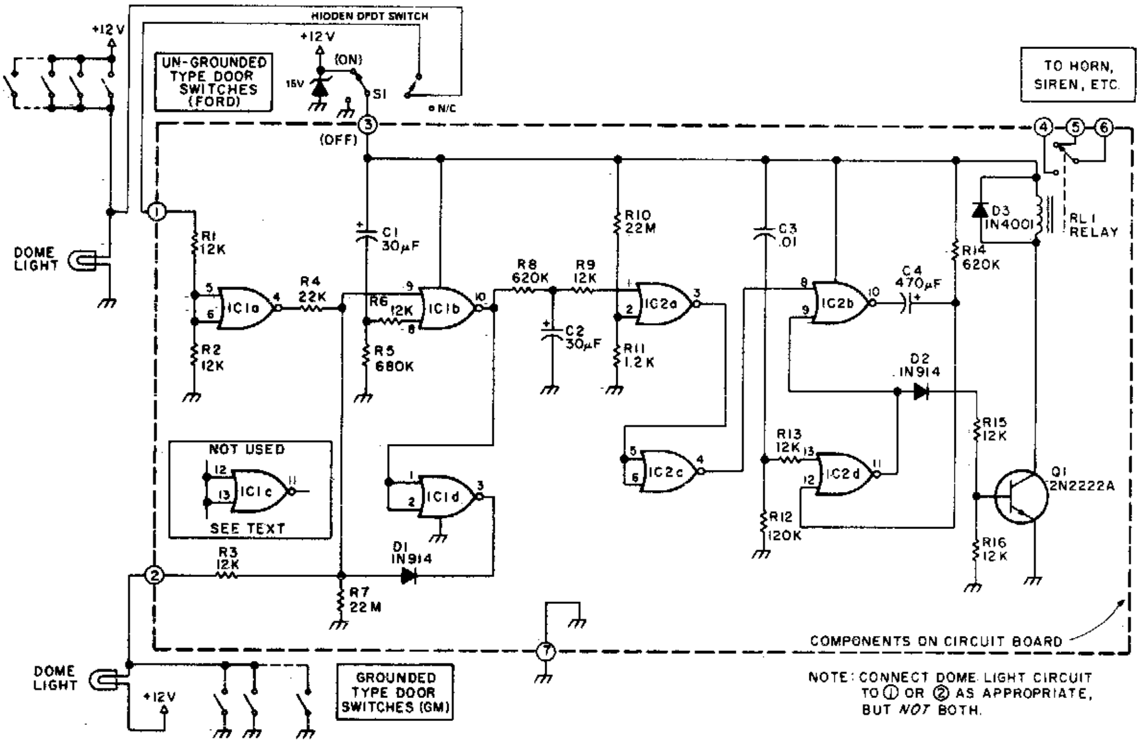


Fig. 1-4

Circuit Notes

This alarm gives a 15-20 second exit and entrance delay. After being triggered, the alarm sounds for five minutes and then shuts off. Once triggered, the sequence is automatic and is not affected by subsequent opening or closing of doors.

HOME SECURITY MONITOR SYSTEM

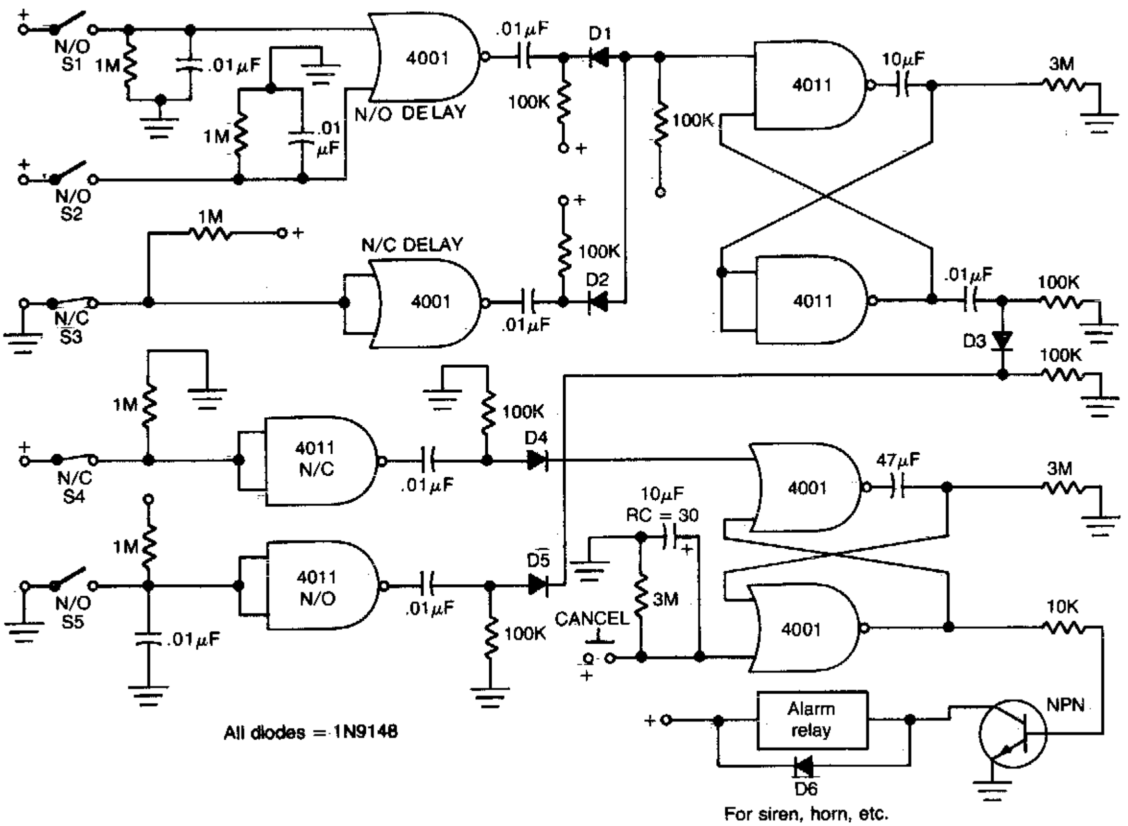


Fig. 1-5

Circuit Notes

This circuit provides normally open (NO) and normally closed (NC) contacts S1, S2, and S3 to turn on the alarm after a 30 second delay. S4 and S5 operate instantly. The CANCEL switch resets the alarm.

ANTITHEFT DEVICE

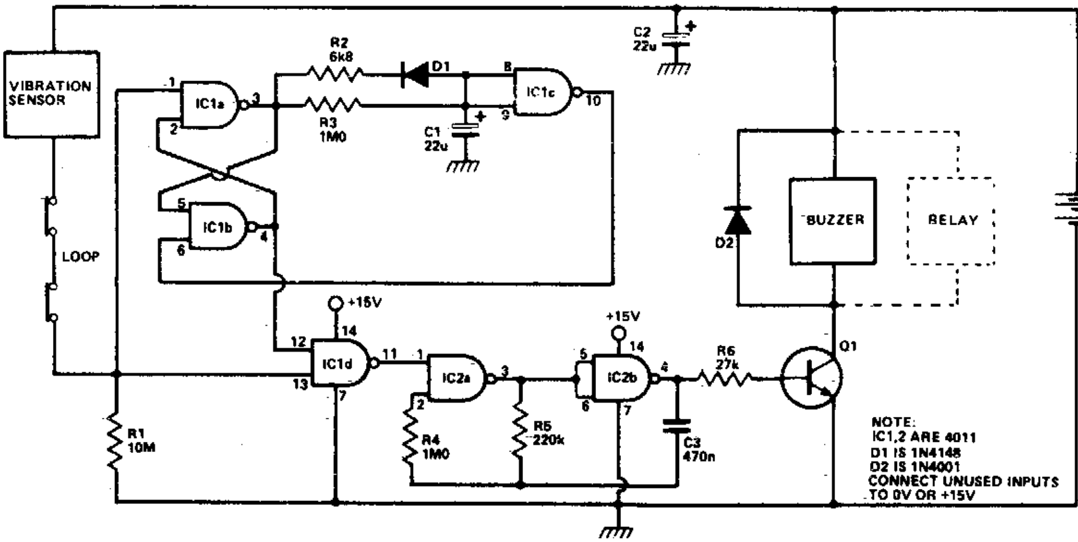


Fig. 1-6

Circuit Notes

Any momentary break in the protective loop or tripping of the normally closed vibration sensor, causes alarm to sound for 20 seconds. If the circuit is open all the time, the alarm will sound continuously.

AUTO BURGLAR ALARM

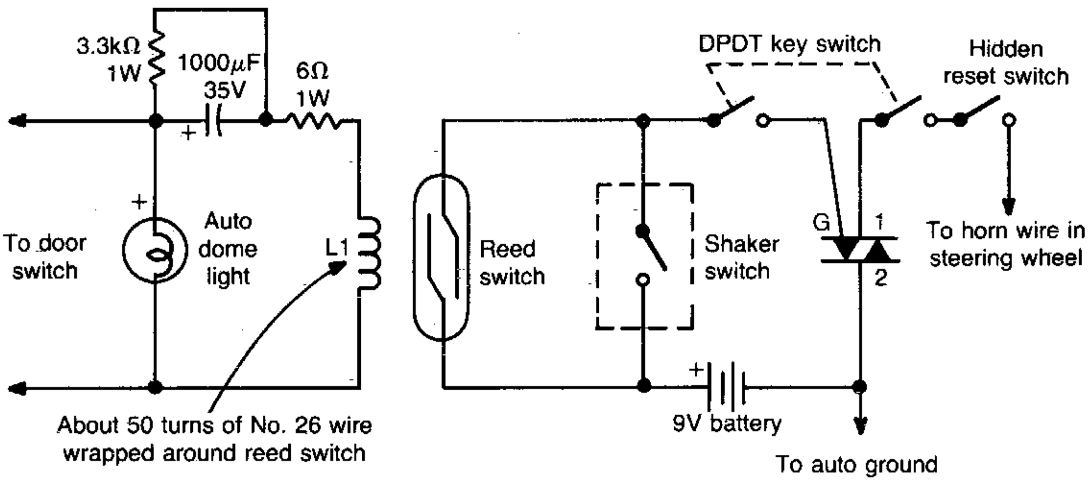


Fig. 1-7

Circuit Notes

Dome light current through L1 closes reed switch and sounds alarm. Shaker switch also activates alarm.

TAMPER-PROOF BURGLAR ALARM

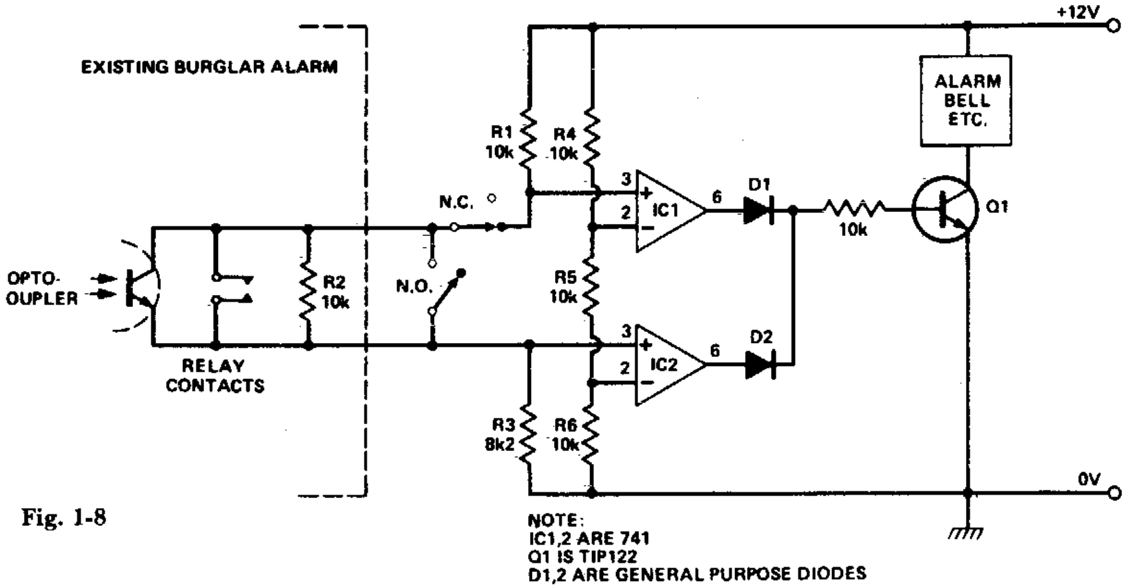


Fig. 1-8

Circuit Notes

If R2 is opened or shorted, the alarm sounds.

LATCHING BURGLAR ALARM

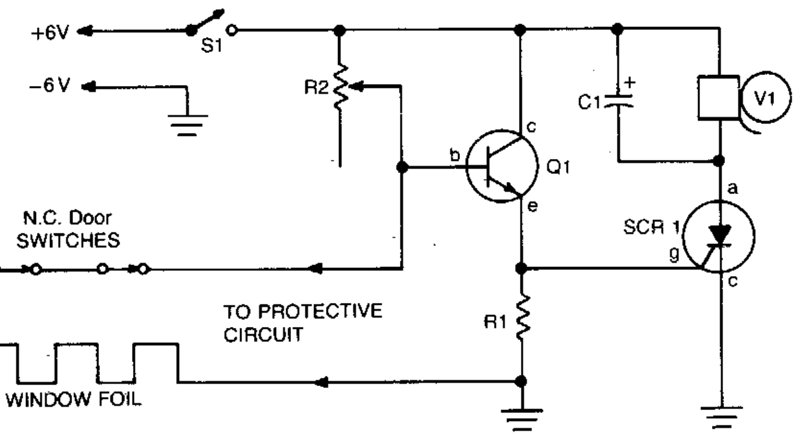


Fig. 1-9

Circuit Notes

When the protective circuit is interrupted (opened), the alarm sounds. To set the circuit, adjust R2 (with protective circuit open) for 1 V across R1.

MOTION-ACTIVATED MOTORCYCLE OR CAR ALARM

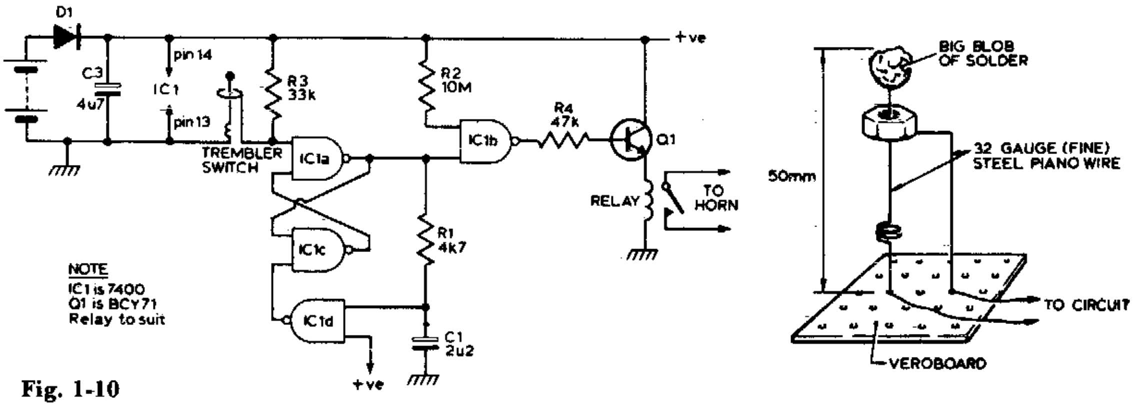


Fig. 1-10

Circuit Notes

Trembler (motion activated) switch sounds the alarm for 5 seconds. Then it goes off. Circuit is timed out for 10 seconds to allow the trembler switch to settle.

BOAT ALARM

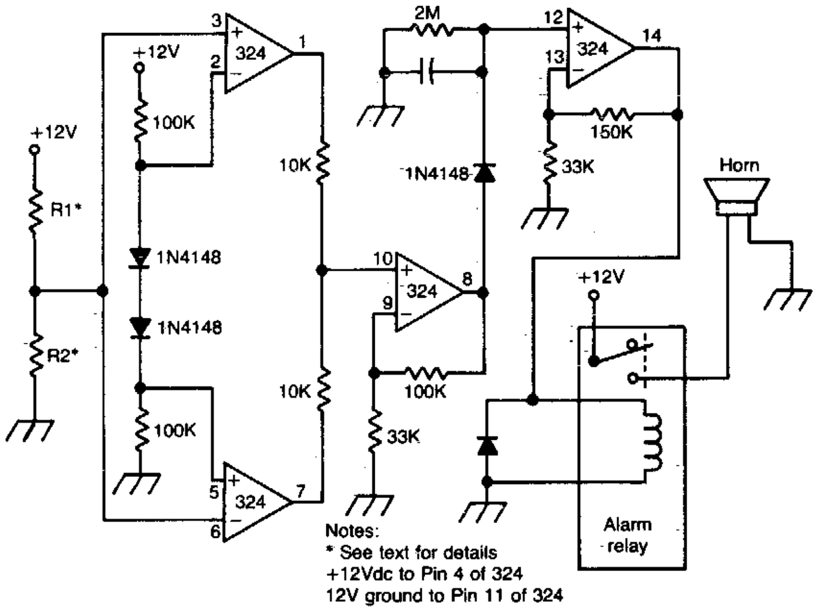


Fig. 1-11

Circuit Notes

Removing R1 or R2 from the circuit (i.e., the potential thief breaks a hidden wire that connects R1 to +12 V and R2 to ground) activates the alarm for about five minutes.

BLOWN-FUSE ALARM

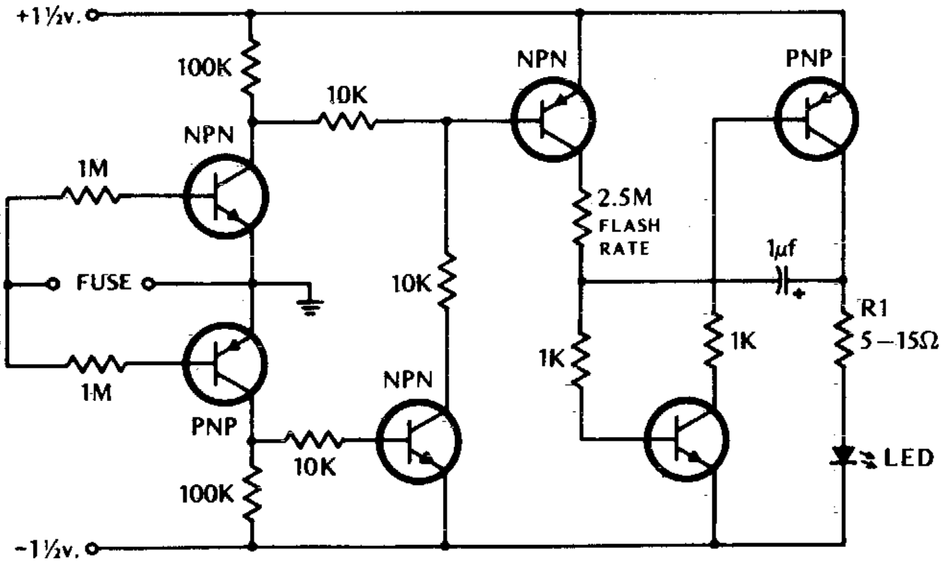


Fig. 1-12

Circuit Notes

If the fuse blows, the LED indicator starts to blink.

AUTO BURGLAR ALARM

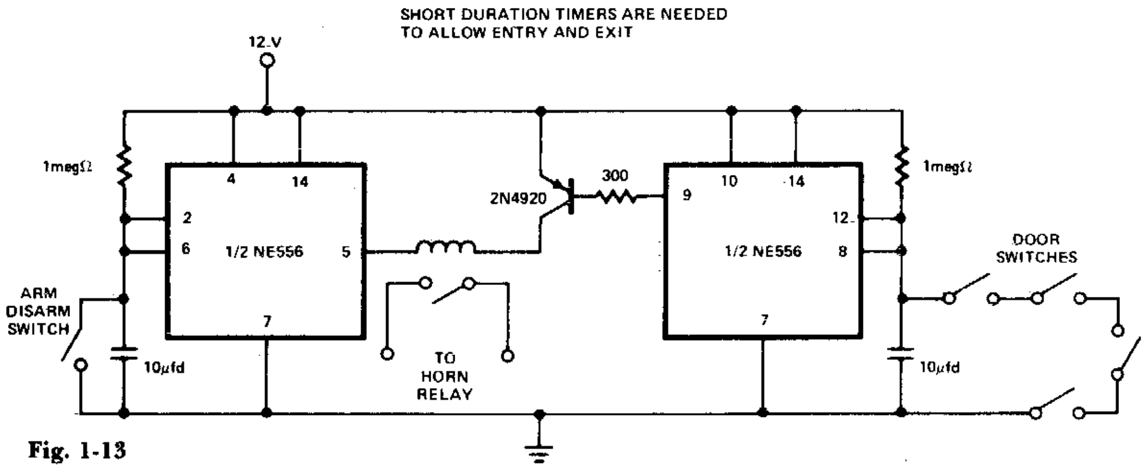


Fig. 1-13

CONTINUOUS-TONE 2 kHz BUZZER WITH BRIDGE DRIVE, GATED ON BY A LOGIC 0

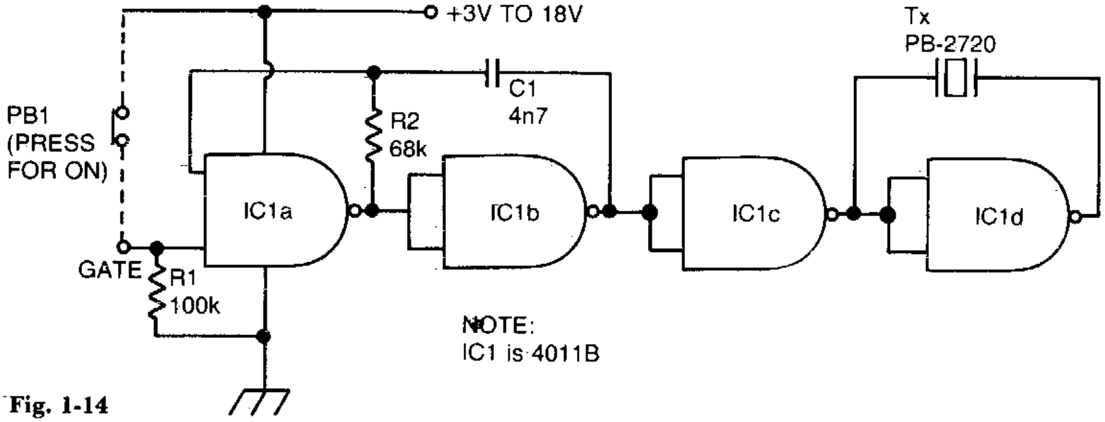


Fig. 1-14

PULSED-TONE ALARM, GATED BY A HIGH INPUT, WITH DIRECT-DRIVE OUTPUT

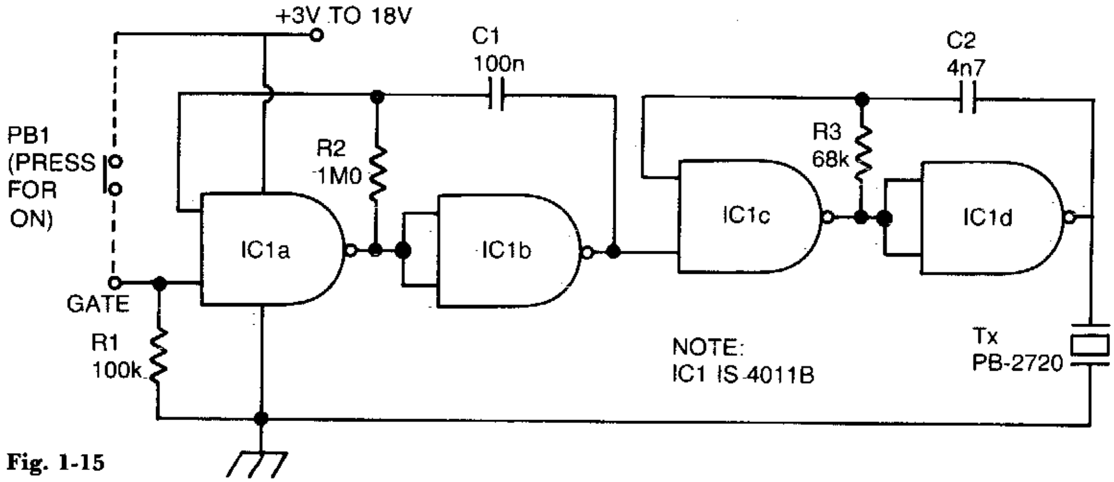


Fig. 1-15

PIEZOELECTRIC ALARM

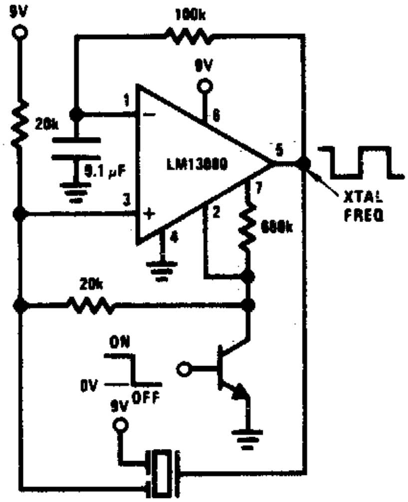


Fig. 1-16

BURGLAR ALARM

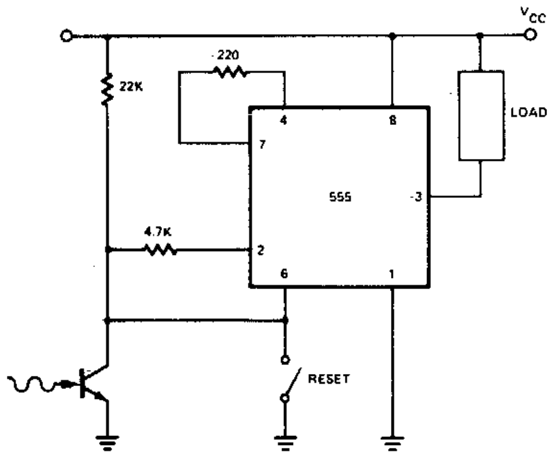


Fig. 1-18

GATED 2 kHz BUZZER

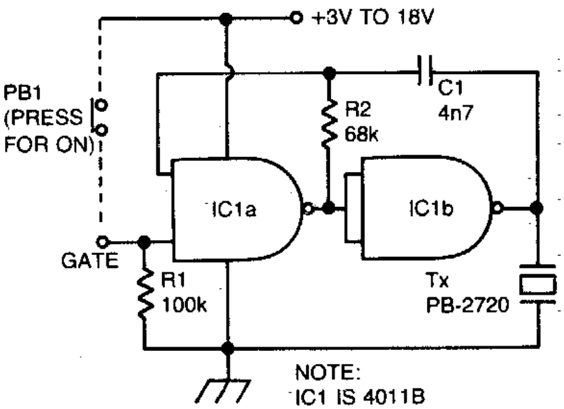


Fig. 1-17

LATCHING BURGLAR ALARM

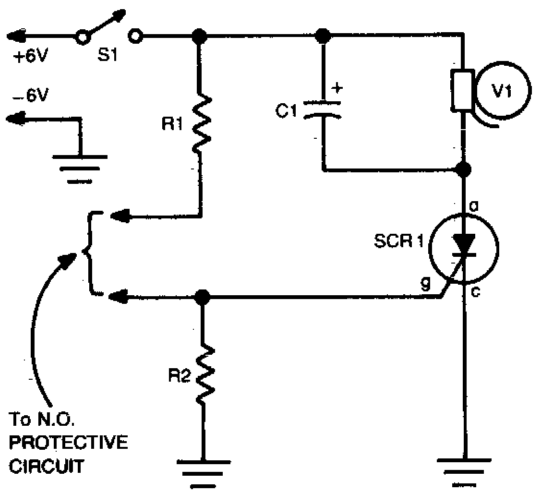


Fig. 1-19

Circuit Notes

Closing the protective circuit (i.e., R1 to R2) applies positive voltage to the gate of SCR1 and sounds the alarm. It can only be turned off with S1.

SUN-POWERED ALARM

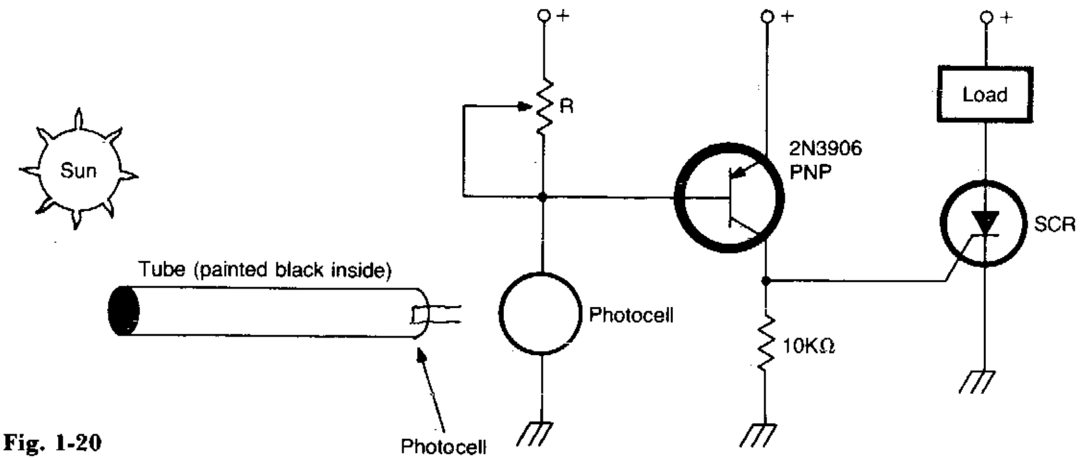


Fig. 1-20

Circuit Notes

Circuit turns on when light (sunlight) strikes photocell. Potentiometer R sets light level at which the alarm sounds. Painted tube (black on inside) may be used on photocell to aim at the sun.

FREEZER MELTDOWN ALARM

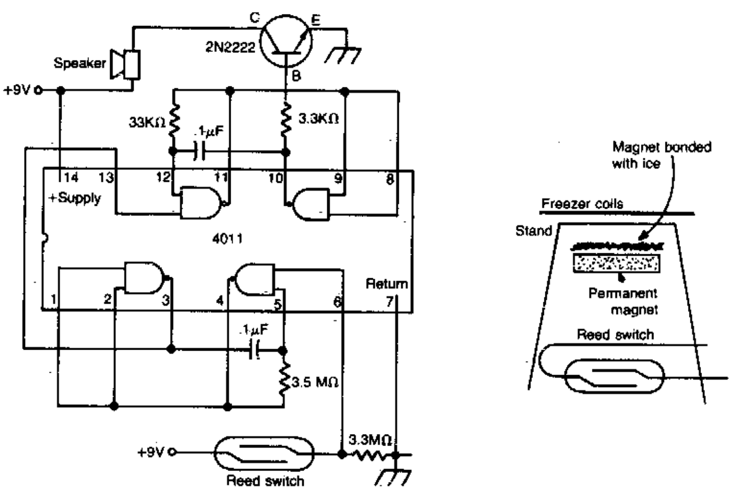


Fig. 1-21

Circuit Notes

The meltdown is a magnet held to a small stand by ice. A reed switch is below the magnet. When the ice melts, the magnet falls on the switch, closing it, and completing the alarm circuit.

2
Amateur Radio

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

CODE-PRACTICE OSCILLATOR PRODUCES AUTOMATIC DITS AND DAHS

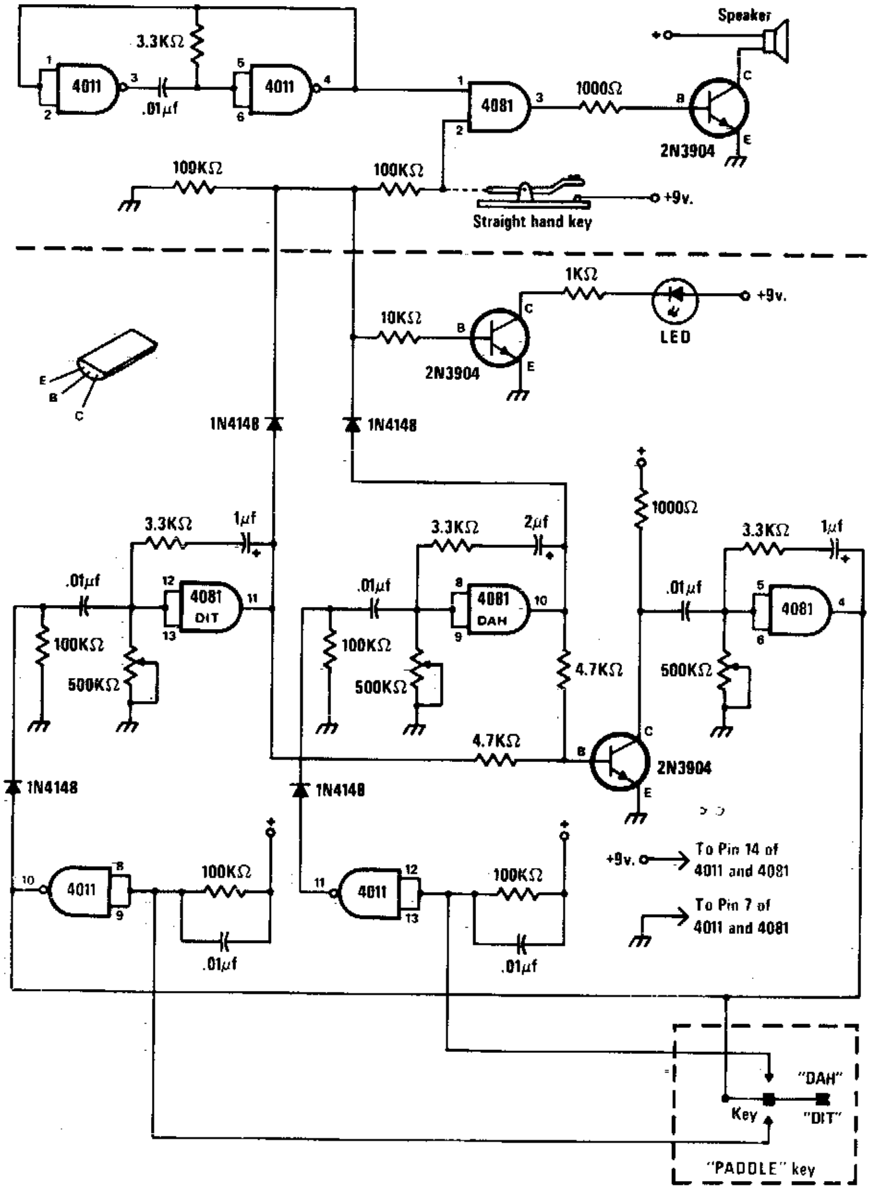


Fig. 2-1

Circuit Notes

The circuit consists of a basic oscillator (above dashed line) and an automatic keyer (below dashed line). The unit can be used with a straight hand key or a paddle key for automatic operation.

RF POWER METER

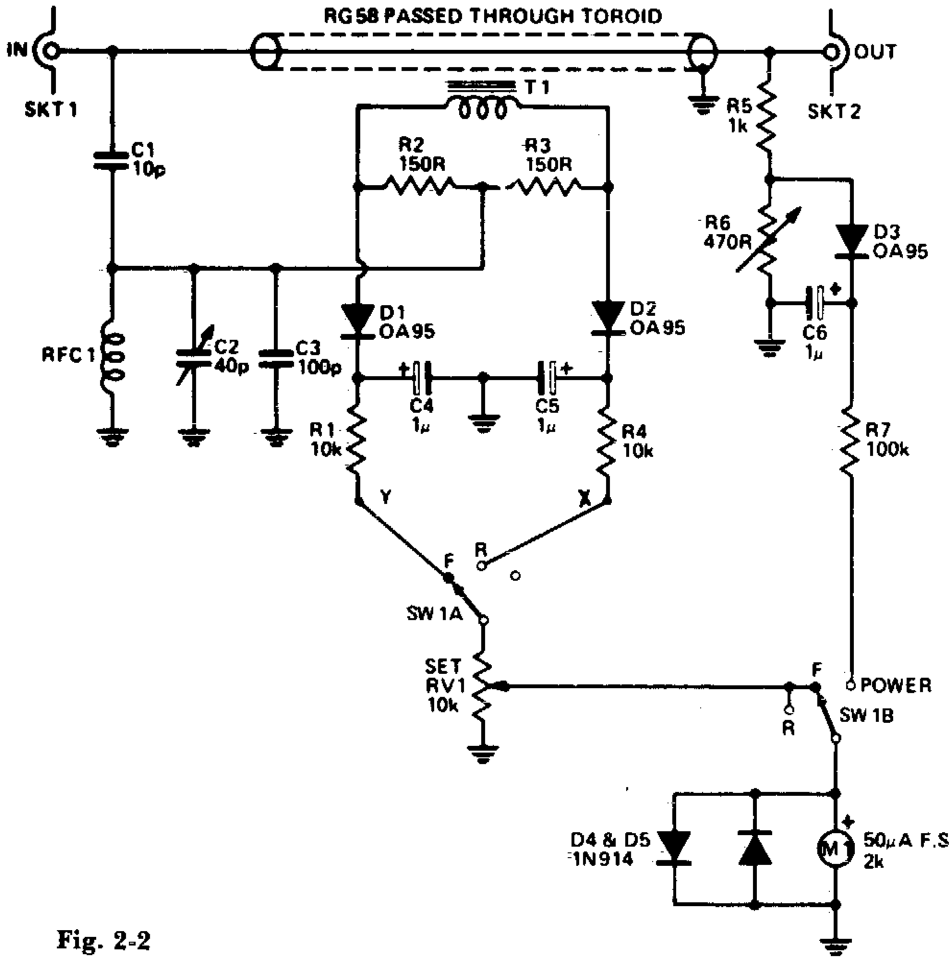


Fig. 2-2

Circuit Notes

Reflectometer (SWR Power Meter) covers three decades—from 100 kHz to 100 MHz. It can be constructed for rf powers as low as 500 mW or up to 500 watts.

IN-LINE WATTMETER

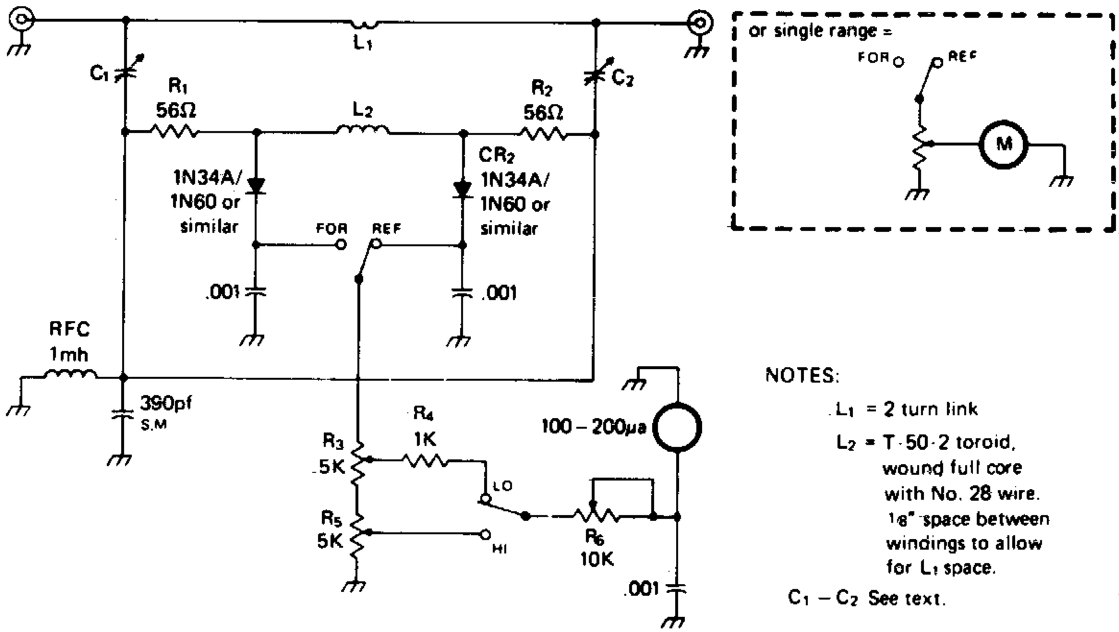


Fig. 2-3

Circuit Notes

The circuit is not frequency sensitive. Its calibration will be accurate over a wide frequency spectrum, such as the entire amateur hf spectrum, if the values of L2, the voltage divider capacitors C1-2 and C3, and the resistances of R1-2 are chosen properly. R1-2 and CR1-2 should be matched for best results. Generally, R1-2 must be small compared to the reactance of L2 so as to avoid any significant effect on the L2 current which is induced by the transmission line current flowing through L1. The lower frequency limit of the bridge is set by the R1-R2/Ls ratio, and the cutoff is at the point where the value of R1-R2 becomes significant with reference to the reactance of L2 at that frequency point.

CW SIGNAL PROCESSOR

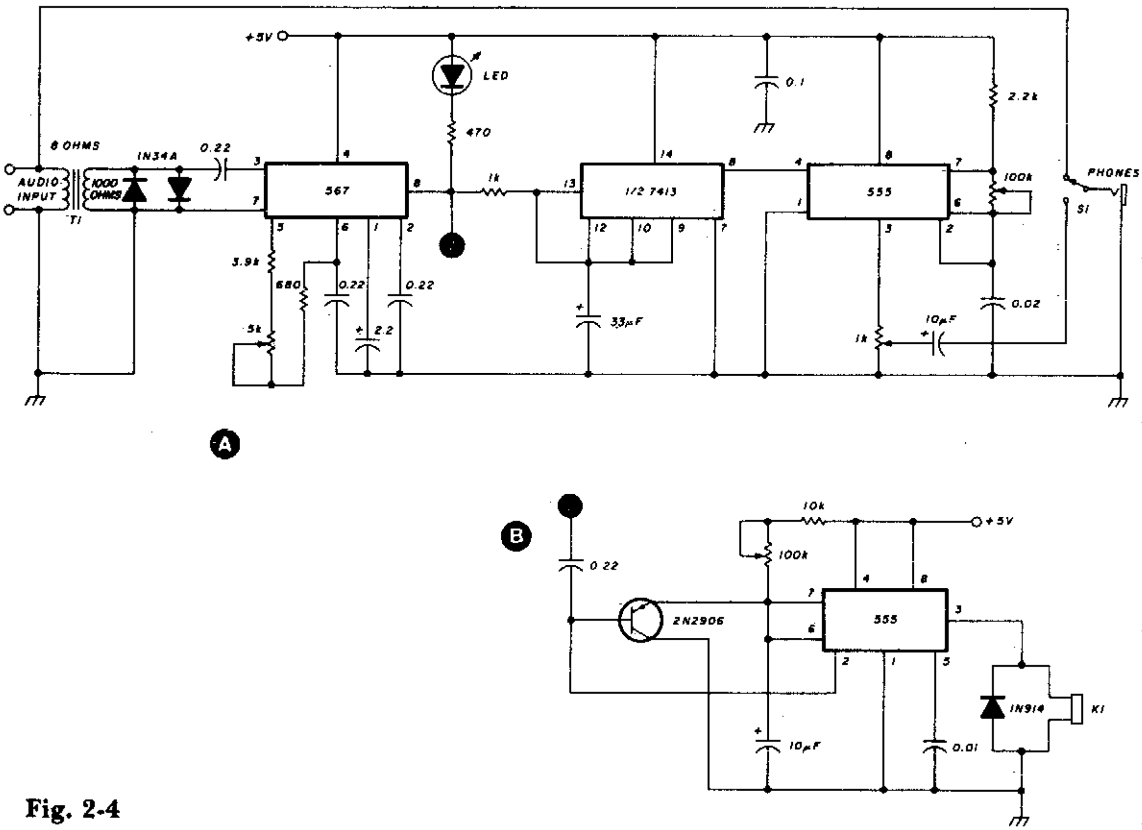
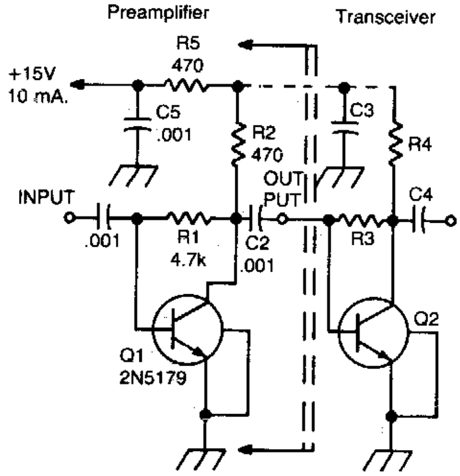


Fig. 2-4

Circuit Notes

This circuit provides interferenced rejection for the CW operator. The 567 phase-locked loop is configured to respond to tones from 500 to 1100 Hz. The Schmitt trigger reduces the weighting effect caused by the output of the PLL remaining low after removal of the audio signal. Ten to 15 millivolts of audio activate the circuit. For periods of loss of signal, circuit B will automatically switch back to live receiver audio after a suitable delay. (If a relay with a 5-volt coil is not available, the circuit can also be powered from +12 volts.) When circuit B is used, the contacts on relay K1 replace S1.

TWO-METER PREAMPLIFIER FOR HANDITALKIES



Circuit Notes

This simple, inexpensive, wideband rf amplifier provides 14 dB gain on two meters without the use of tuned circuits.

Fig. 2-5

REPEATER BEEPER

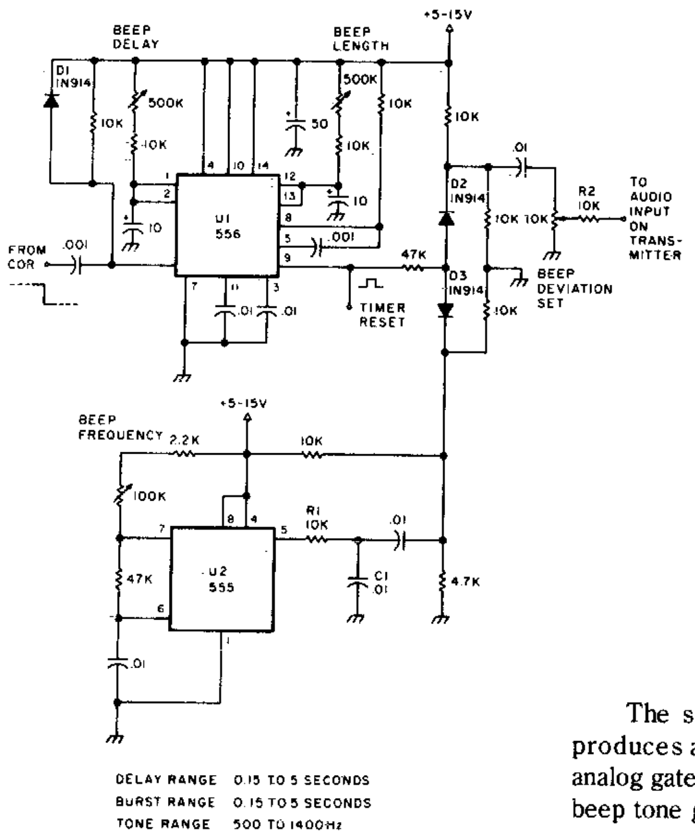


Fig. 2-6

Circuit Notes

The signal from COR triggers U1 which produces a beep-gate pulse that enables the analog gate consisting of D2 and D3 to pass the beep tone generated by U2.

ELECTRONIC KEYER

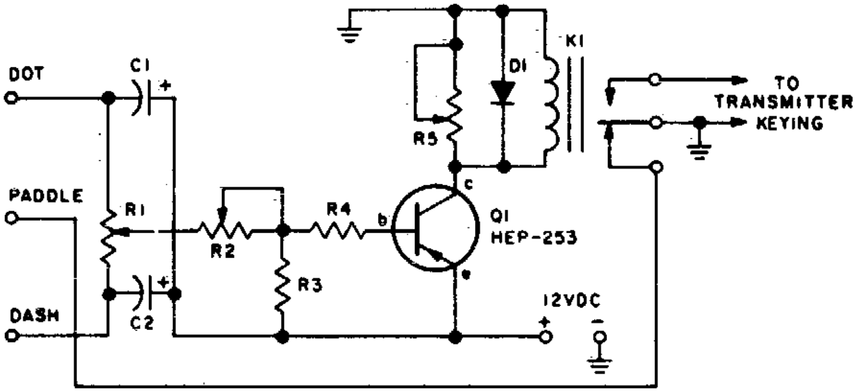


Fig. 2-7

PARTS LIST FOR HAM'S KEYER

C1—3-uF, 6-VDC electrolytic capacitor
C2—10-uF, 6-VDC electrolytic capacitor
D1—1N60 diode
K1—12-VDC relay
Q1—HEP-253 pnp transistor
R1—10,000-ohm linear potentiometer
R2—50,000-ohm potentiometer
R3—1200-ohm, ½-watt resistor
R4—560-ohm, ½-watt resistor
R5—5000-ohm potentiometer

Circuit Notes

This circuit automatically produces Morse code dots and dashes set by time constants involving C1 and C2. R1 sets dot/dash ratio and R2 sets the speed. R5 sets the relay drop-out point.

CODE PRACTICE OSCILLATOR

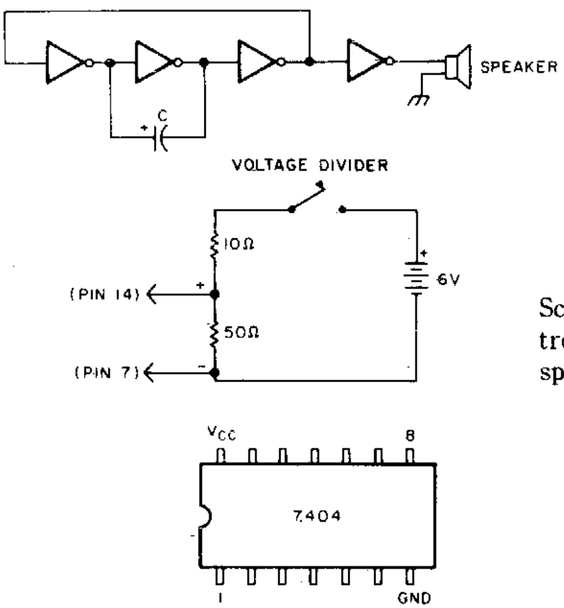


Fig. 2-8

Circuit Notes

This simple cpo uses the 7404 low-power Schottky hex inverter. C is a 5- to 30-μF electrolytic selected for the desired pitch. The speaker is a 2-inch, 8-ohm unit.

AUTOMATIC TAPE RECORDING

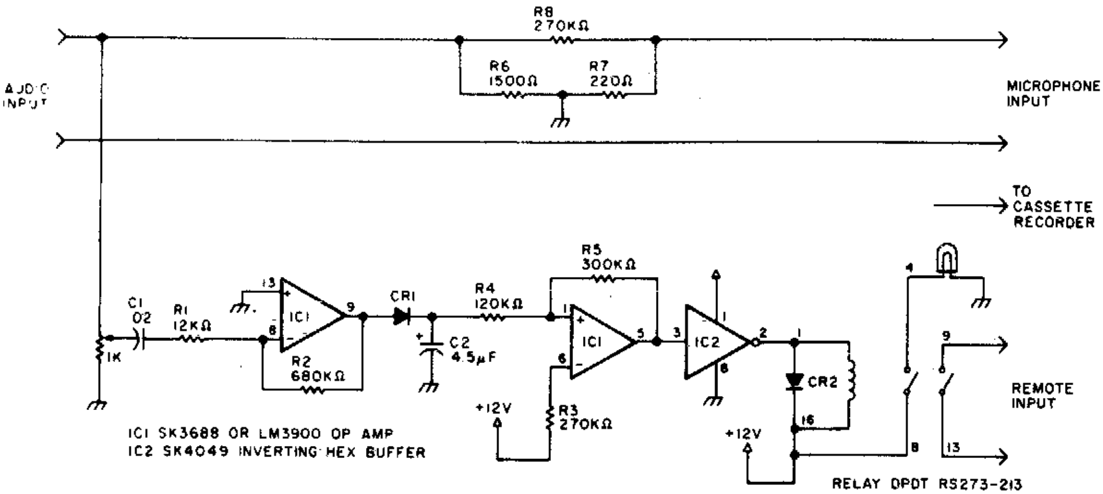


Fig. 2-9

Circuit Notes

Amateurs don't have to miss the action while away from the rig. This circuit turns on a tape recorder whenever the receiver's squelch is broken. After signal loss, the recorder will shut off following a slight delay.

SELF-POWERED CW MONITOR

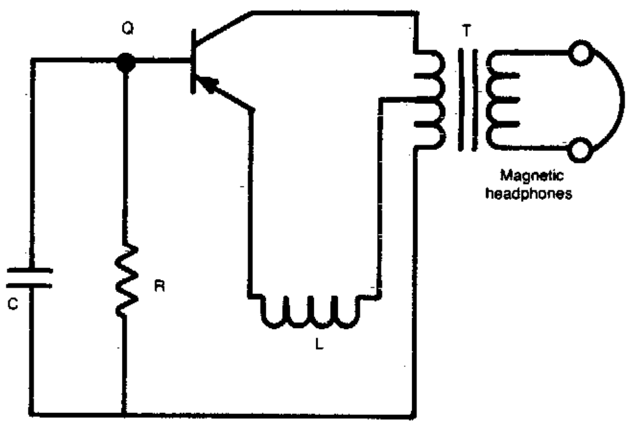


Fig. 2-10

Circuit Notes

Position L near the transmitter output tank to hear the key-down tone. Then tape the coil in place. C = .047 μF, R = 8.2 K, Q = HEP 253 (or equal), T = 500: 500 ohm center tapped transformer. L = 2 to 6 turns on ½" coil form.

REMOTE RF CURRENT READOUT

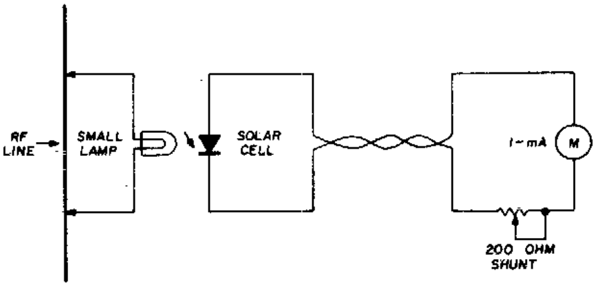


Fig. 2-11

Circuit Notes

A suitable pilot lamp is illuminated by a small sample of rf and energizes an inexpensive solar cell; the dc current generated by the cell is a measure of relative rf power, and may be routed to a low-current meter located at any convenient point. A sensitive, low-current pilot lamp is desirable to cause minimum disturbance to normal rf circuit conditions. The number 48 or 49, 60 mA lamp is suitable for use with transmitters above 1-watt output.

CODE PRACTICE OSCILLATOR

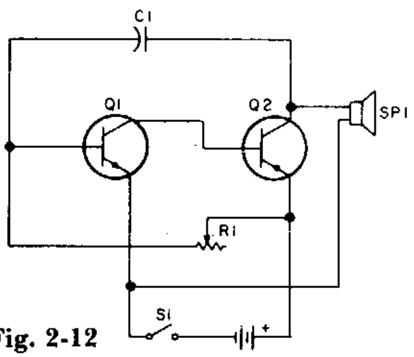


Fig. 2-12

Circuit Notes

Oscillator, works with 2 to 12 Vdc (but 9 to 12 volts gives best volume and clean keying). R1 can be replaced with a 500 K pot and the circuit will sweep the entire audio frequency range.

SWR WARNING INDICATOR

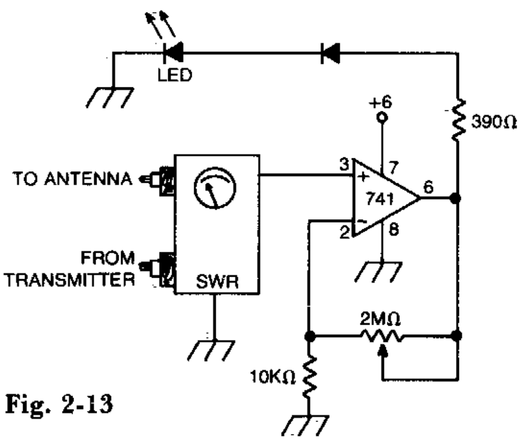


Fig. 2-13

Circuit Notes

Op amp with dc input from SWR meter can be adjusted to preset the SWR reading at which the LED lights.

SUBAUDIBLE TONE ENCODER

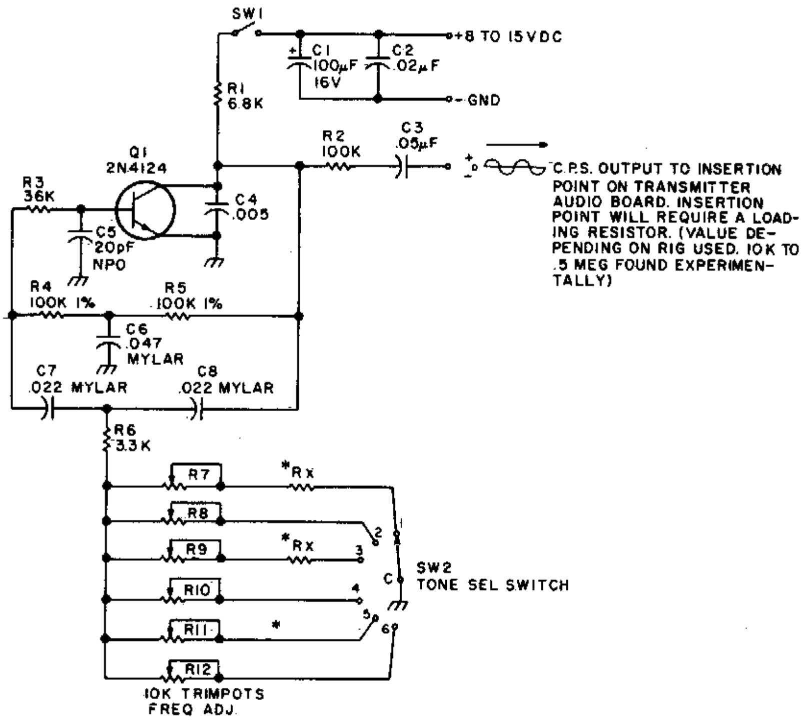


Fig. 2-14

Circuit Notes

This twin-T oscillator produces six preset subaudible tones from 93 to 170 Hz in three ranges.

AUDIO MIXER

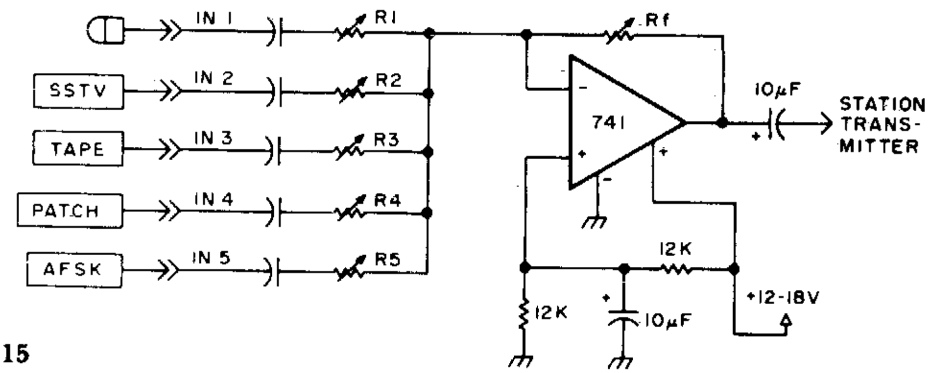
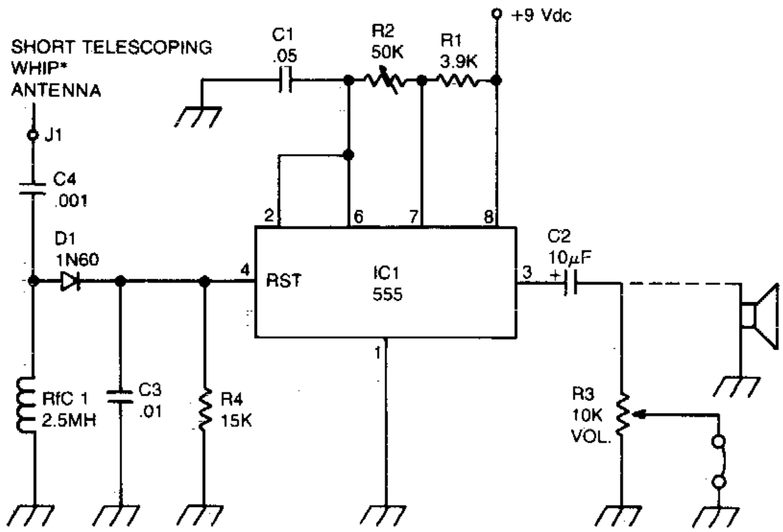


Fig. 2-15

Circuit Notes

The 741 op amp is used as a summing amplifier to combine several audio inputs. Overall gain is set by R_f.

RF-POWERED SIDETONE OSCILLATOR



*PORTABLE RADIO REPLACEMENT TYPE

Fig. 2-16

Circuit Notes

A sidetone oscillator is a special audio oscillator that is turned on and off with the transmitter. The oscillator is rf-driven and battery operated. It uses a 555 IC timer as an astable multivibrator. Keying is accomplished by applying a positive dc potential, developed from the rf signal, to the reset terminal of the 555.

HARMONIC GENERATOR

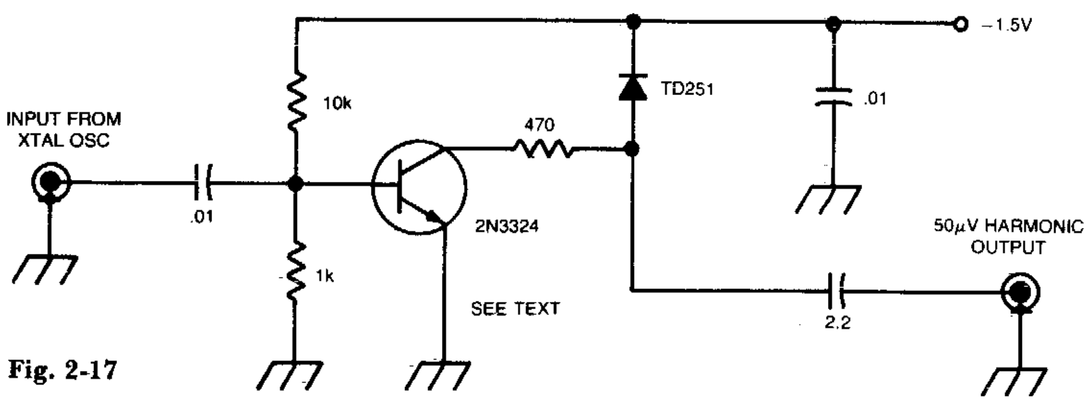


Fig. 2-17

Circuit Notes

This circuit will produce 50 μV harmonics through 1296 MHz with an input of 0.15-1 V from a 100 or 1000 kHz crystal oscillator. With a germanium diode instead of a tunnel diode, harmonics can be heard up to about 147 MHz.

AUTOMATIC TTL MORSE-CODE KEYER

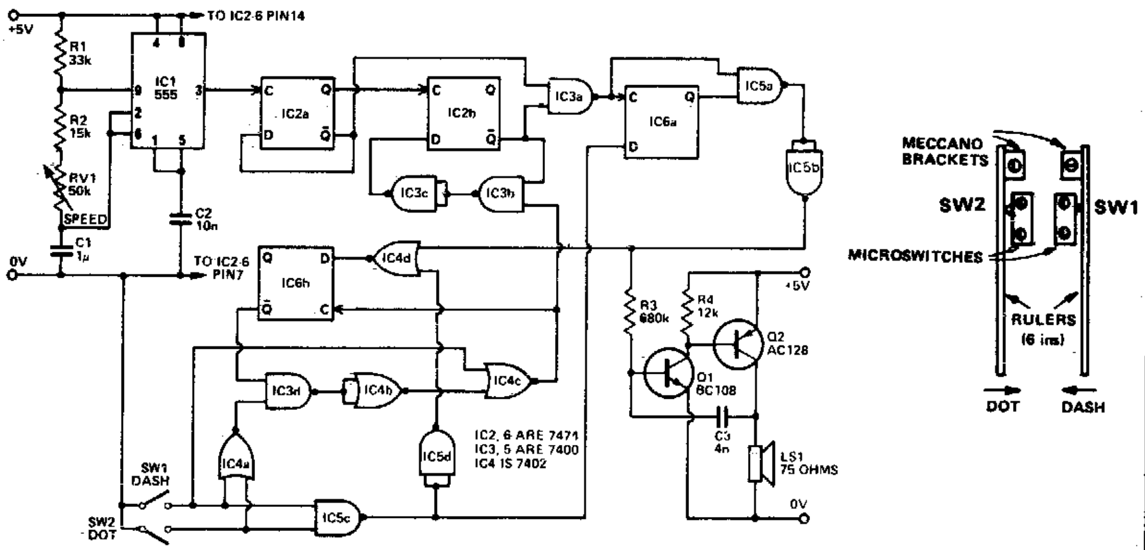


Fig. 2-18

Circuit Notes

Automatically generated dits and dahs are produced over a speed range of 11 to 39 wpm. The upper limit can be raised by decreasing R2. SW1 and SW2 can be a "home-brew" paddle operated key.

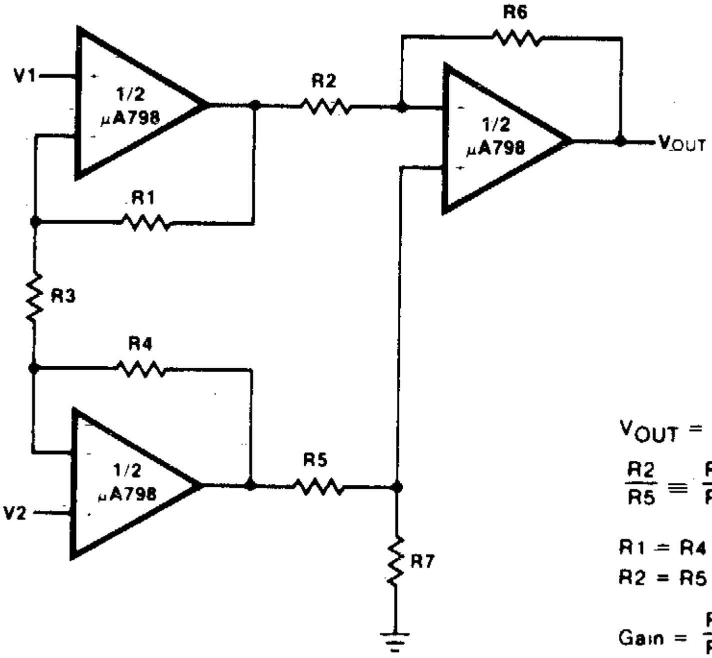
3

Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

High Impedance Differential Amplifier
Unity Gain Follower
Voltage Controlled Variable Gain Amplifier
Power Booster
Logarithmic Amplifier
Voltage Controlled Variable Gain Amplifier
Discrete Current Booster
Precision Process Control Interface
Voltage Controlled Amplifier
Absolute Value Amplifier
Programmable Gain Noninverting Amplifier with Selectable Inputs
× 1000 Amplifier Circuit
Inverting Amplifier with Balancing Circuit
Switching Power Amplifier
Precision Power Booster
Noninverting Voltage Follower
Color Video Amplifier
Fast Voltage Follower
Isolation Amplifier for Capacitive Loads
Cable Bootstrapping
Current Booster
Wideband Unity Gain Inverting Amplifier in a 75 Ohm System
High-Speed Current to Voltage Output Amplifier
Gated Amplifier
Reference Voltage Amplifier
Fast Summing Amplifier
Adjustment-Free Precision Summing Amplifier
Summing Amplifier with Low Input Current
× 10 Operational Amplifier Using L161
× 100 Operational Amplifier Using L161
Precision Absolute Value Circuit
Ultra-Low-Leakage Preamp
Dc to Video Log Amplifier
±100 V Common Mode Range Differential Amplifier
Wide Bandwidth, Low Noise, Low Drift Amplifier
Signal Distribution Amplifier
Audio Distribution Amplifier
High Input Impedance, High Output Current Voltage Follower
Precision Amplifier
Preamplifier and High-to-Low Impedance Converter
Noninverting Amplifier
High Impedance, High Gain, High Frequency Inverting Amp
Log-Ratio Amplifier
Inverting Amplifier
Logarithmic Amplifier

HIGH IMPEDANCE DIFFERENTIAL AMPLIFIER



$$V_{OUT} = C\,(1 + a + b)(V2 - V1)$$

$$\frac{R2}{R5} \equiv \frac{R6}{R7} \text{ for best CMRR}$$

$$R1 = R4$$

$$R2 = R5$$

$$\text{Gain} = \frac{R6}{R2}\left(1 + \frac{2R1}{R3}\right) = C\,(1 + a + b)$$

Fig. 3-1

UNITY GAIN FOLLOWER

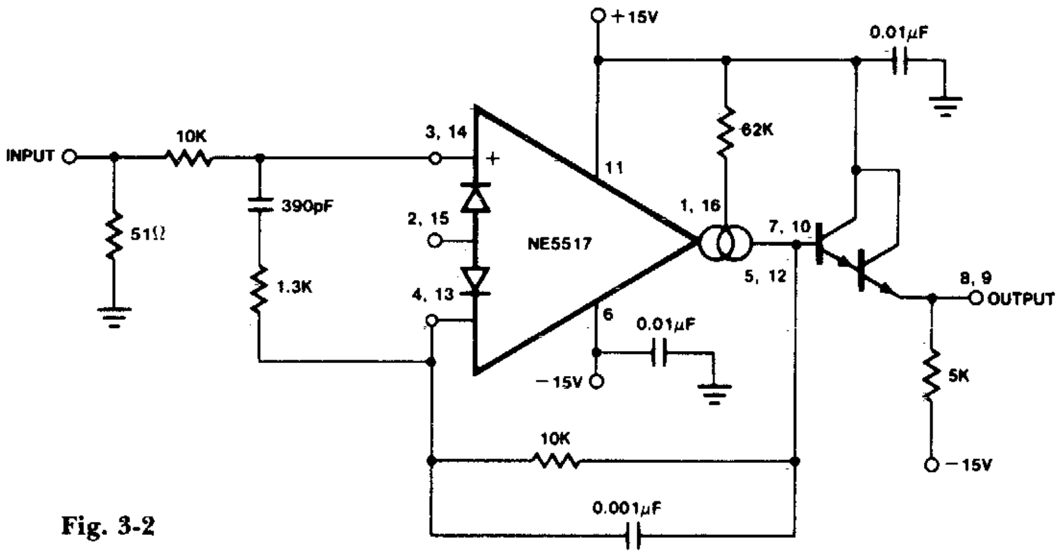


Fig. 3-2

VOLTAGE CONTROLLED VARIABLE GAIN AMPLIFIER

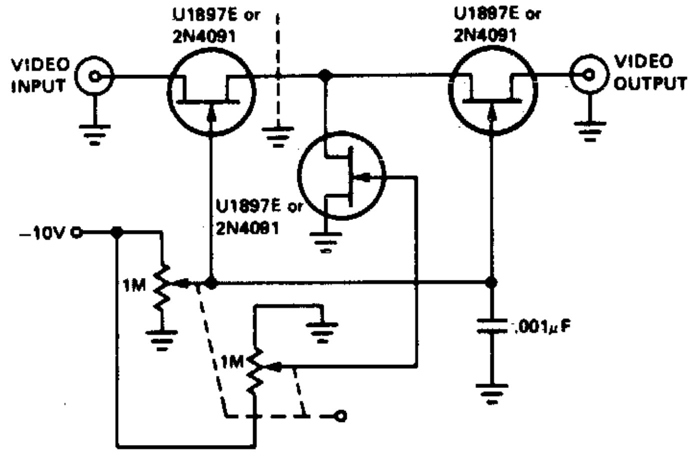
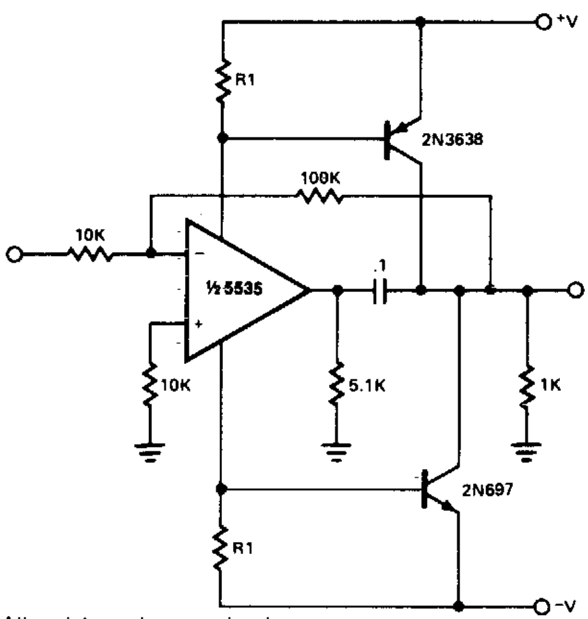


Fig. 3-3

Circuit Notes

The tee attenuator provides for optimum dynamic linear range attenuation up to 100 dB, even at f = 10.7 MHz with proper layout.

POWER BOOSTER



All resistor values are in ohms.

Fig. 3-4

Circuit Notes

Power booster is capable of driving moderate loads. The circuit as shown uses a NE5535 device. Other amplifiers may be substituted only if R1 values are changed because of the Icc current required by the amplifier. R1 should be calculated from the following expression:

$$R1 = \frac{600\ mW}{I_{CC}}$$

LOGARITHMIC AMPLIFIER

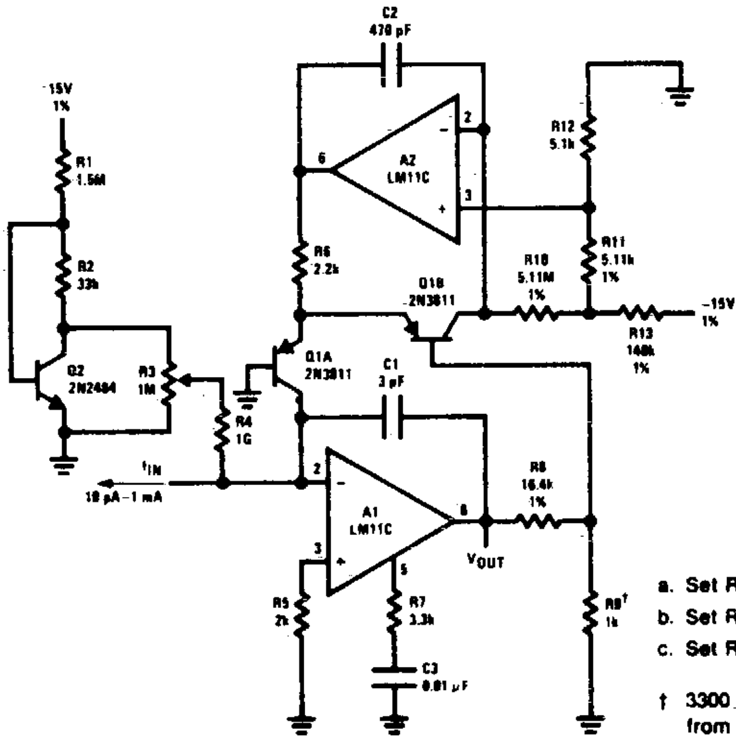


Fig. 3-5

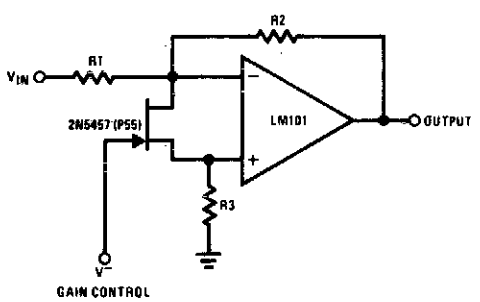
a. Set R11 for $V_{OUT} = 0$ at $I_{IN} = 100\ \mu A$
b. Set R8 for $V_{OUT} = 3V$ at $I_{IN} = 100\ \mu A$
c. Set R3 for $V_{OUT} = -4V$ at $I_{IN} = 10\ pA$

† 3300 ppM/°C. Type Q209 available from Tel Labs, Inc., Manchester, N.H.

Circuit Notes

Unusual frequency compensation gives this logarithmic converter a 100 μs time constant from 1 mA down to 100 μA, increasing from 200 μs to 200 ms from 10 nA to 10 pA. Optional bias current compensation can give 10 pA resolution from –55 °C to 100 °C. Scale factor is 1 V/decade and temperature compensated.

VOLTAGE CONTROLLED VARIABLE GAIN AMPLIFIER



Circuit Notes

The 2N5457 acts as a voltage variable resistor with an $R_{ds(on)}$ of 800 ohms max. Since the differential voltage on the LM101 is in the low mV range, the 2N5457 JFET will have linear resistance over several decades of resistance providing an excellent electronic gain control.

Fig. 3-6

DISCRETE CURRENT BOOSTER

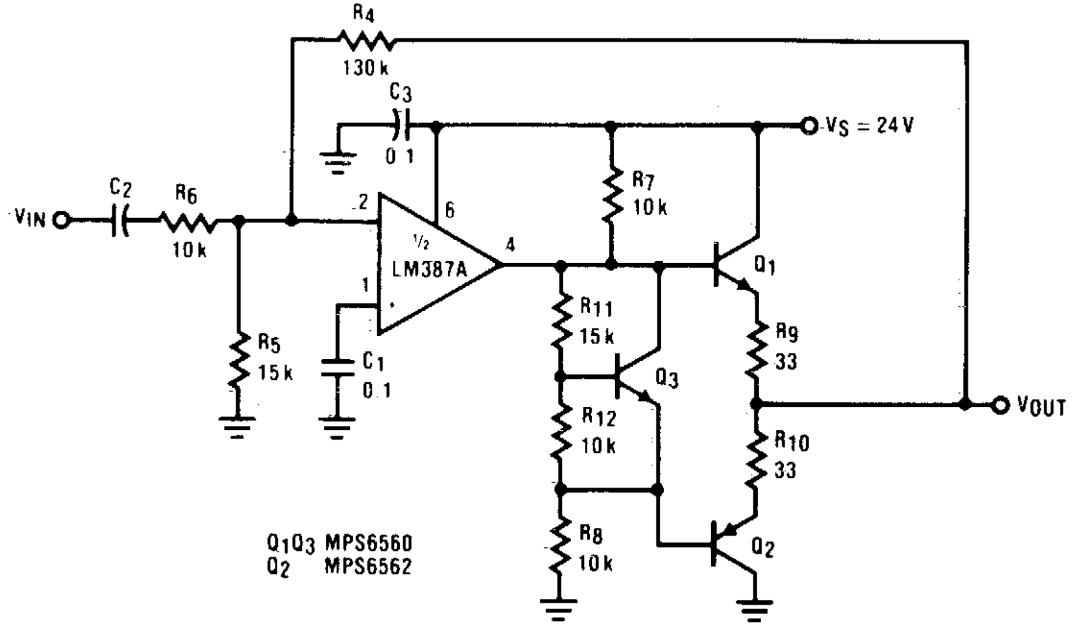


Fig. 3-7

PRECISION PROCESS CONTROL INTERFACE

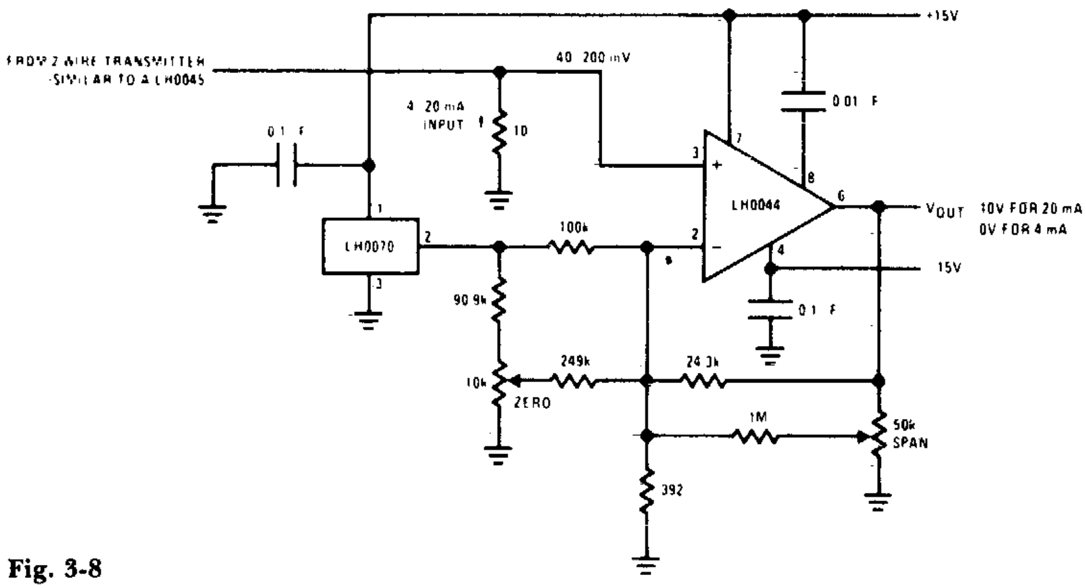


Fig. 3-8

VOLTAGE CONTROLLED AMPLIFIER

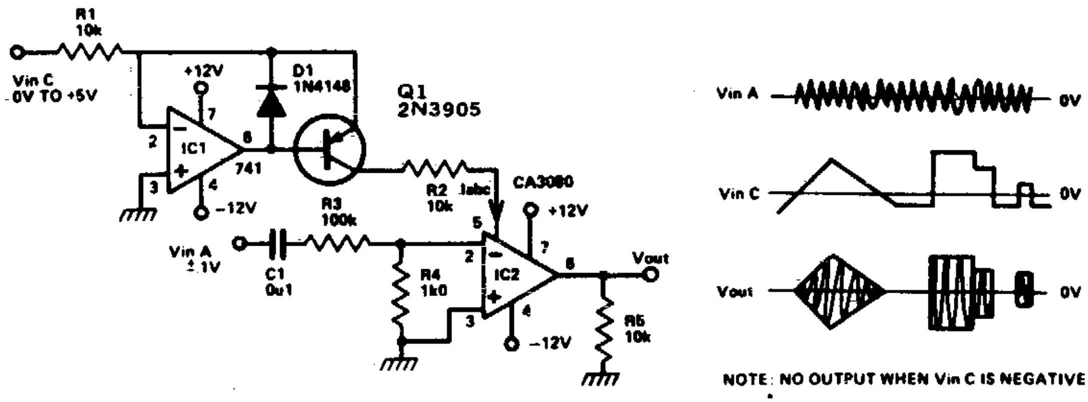


Fig. 3-9

Circuit Notes

This circuit is basically an op amp with an extra input at pin 5. A current I_{ABC} is injected into this input and this controls the gain of the device linerly. Thus by inserting an audio signal (±10 mV) between pin 2 and 3 and by controlling the current on pin 5, the level of the signal output (pin 6) is controlled.

ABSOLUTE VALUE AMPLIFIER

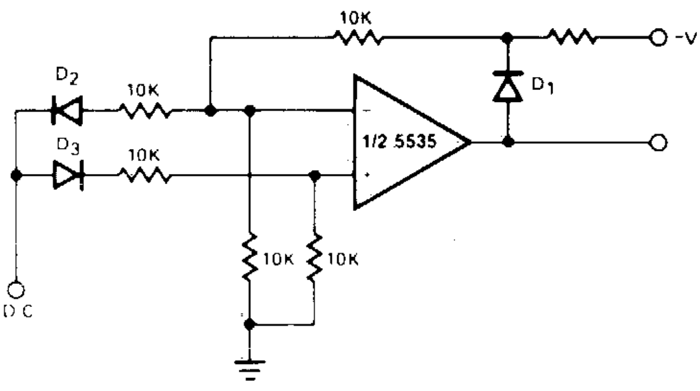


Fig. 3-10

Circuit Notes

The circuit generates a positive output voltage for either polarity of input. For positive signals, it acts as a noninverting amplifier and for negative signals, as an inverting amplifier. The accuracy is poor for input voltages under 1 V, but for less stringent applications, it can be effective.

PROGRAMMABLE GAIN NONINVERTING AMPLIFIER WITH SELECTABLE INPUTS

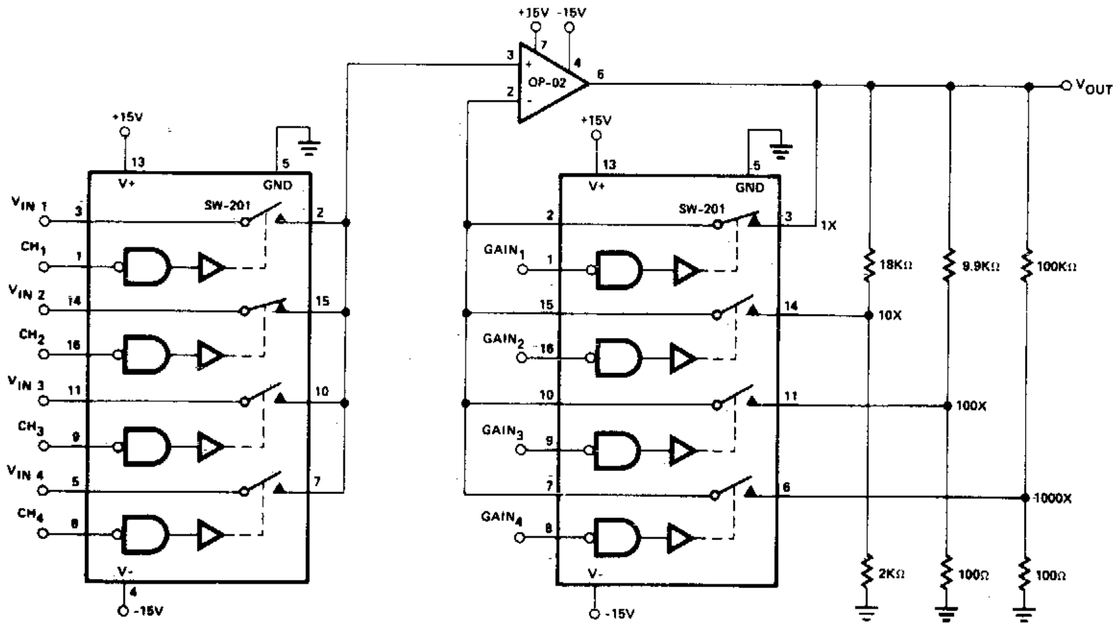


Fig. 3-11

× 1000 AMPLIFIER CIRCUIT

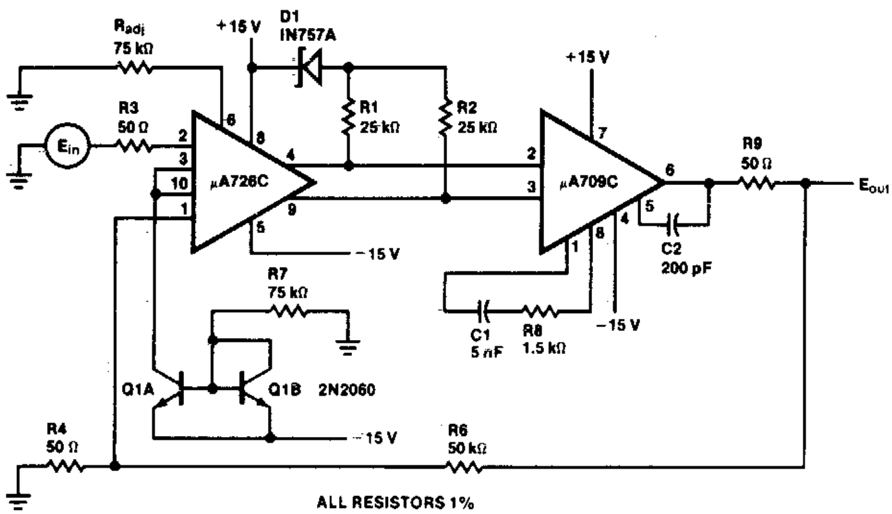


Fig. 3-12

INVERTING AMPLIFIER WITH BALANCING CIRCUIT

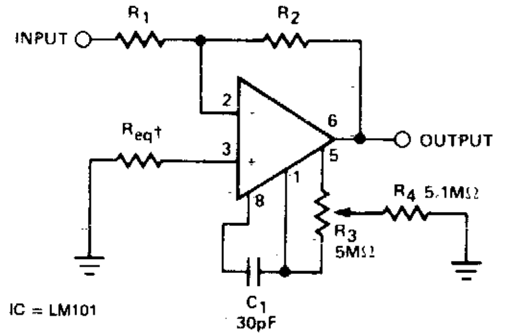


Fig. 3-13

Circuit Notes

R_{eq} may be zero or equal to the parallel combination of R1 and R2 for minimum offset.

PRECISION POWER BOOSTER

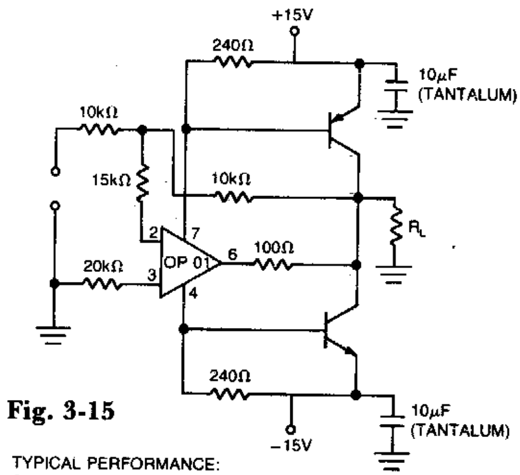


Fig. 3-15

TYPICAL PERFORMANCE:
SLEW RATE------------≈18V/μSEC
0.1% SETTLING-------4μSEC (R_L = 500)
QUIESCENT SUPPLY CURRENT---1.5mA

SWITCHING POWER AMPLIFIER

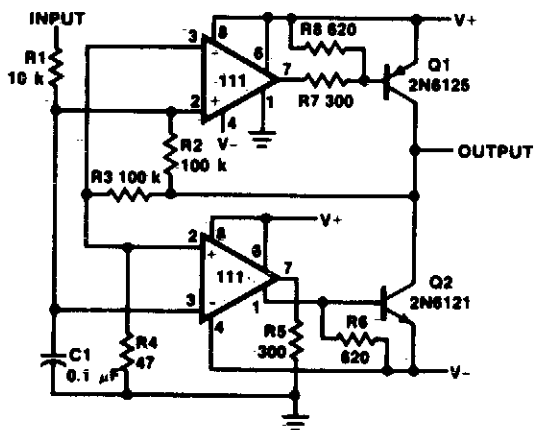


Fig. 3-14

NONINVERTING VOLTAGE FOLLOWER

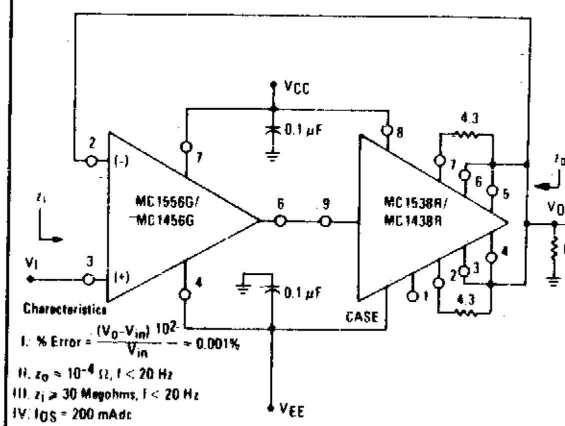
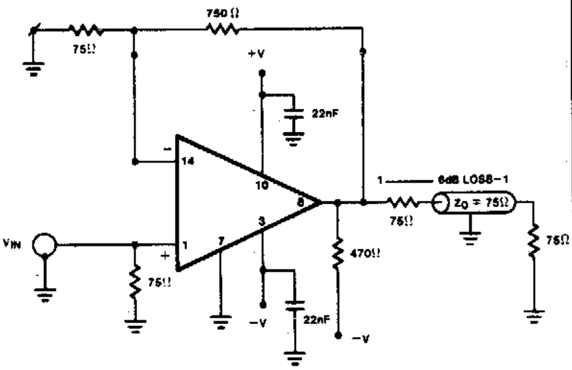


Fig. 3-16

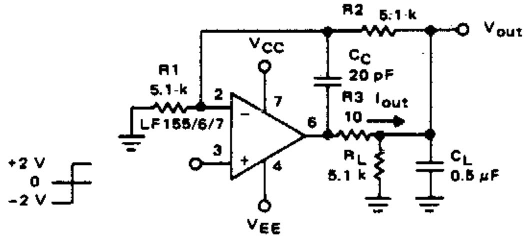
COLOR VIDEO AMPLIFIER



IC = NE5539

Fig. 3-17

ISOLATION AMPLIFIER FOR CAPACITIVE LOADS



- Overshoot 6%
- t_s = 10 μs
- When driving large C_L, the V_{OUT} slew rate is determined by C_L and $I_{out(max)}$:

$$\frac{\Delta V_{out}}{\Delta t} = \frac{I_{out}}{C_L} \cong \frac{0.02}{0.5}\ V/\mu s = 0.04\ V/\mu s \text{ (with } C_L \text{ shown)}$$

Fig. 3-19

FAST VOLTAGE FOLLOWER

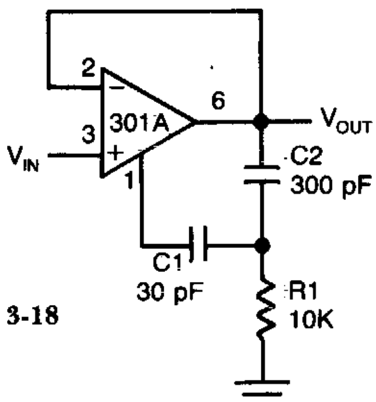


Fig. 3-18

Power Bandwidth: 15 kHz
Slew Rate: 1V/μs

CABLE BOOTSTRAPPING

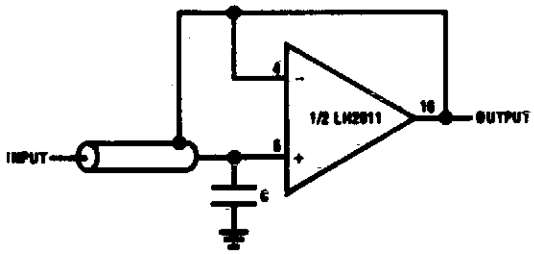
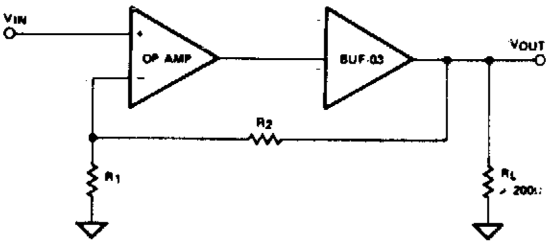


Fig. 3-20

Circuit Notes

Bootstrapping input shield for a follower reduces cable capacitance, leakage, and spurious voltages from cable flexing. Instability can be avoided with small capacitor on input.

CURRENT BOOSTER



$GAIN = \frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_2}{R_1}$

MAXIMUM LOAD CURRENT = ±50mA (10V ÷ 200Ω)

Fig. 3-21

HIGH-SPEED CURRENT TO VOLTAGE OUTPUT AMPLIFIER

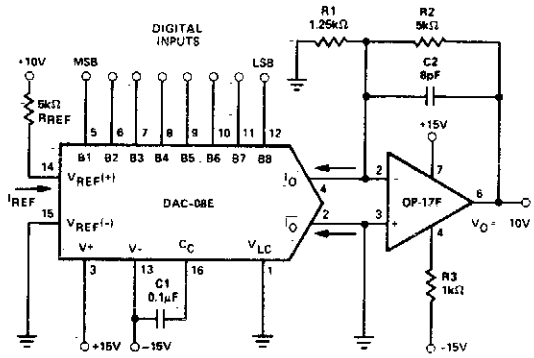


Fig. 3-23

WIDEBAND UNITY GAIN INVERTING AMPLIFIER IN A 75 OHM SYSTEM

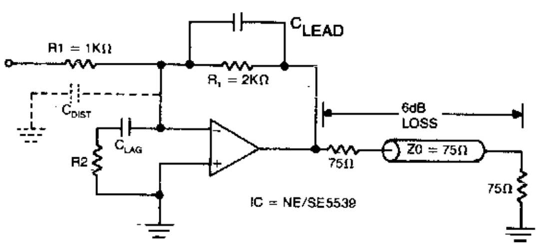


Fig. 3-22

LOGARITHMIC AMPLIFIER

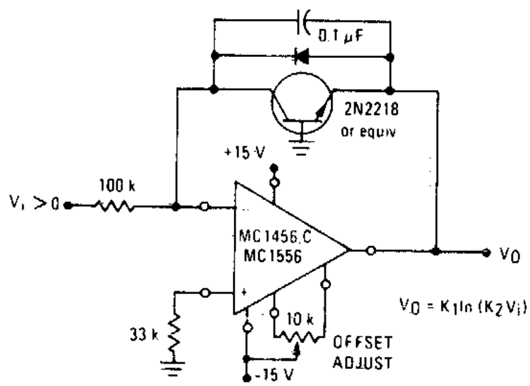


Fig. 3-24

GATED AMPLIFIER

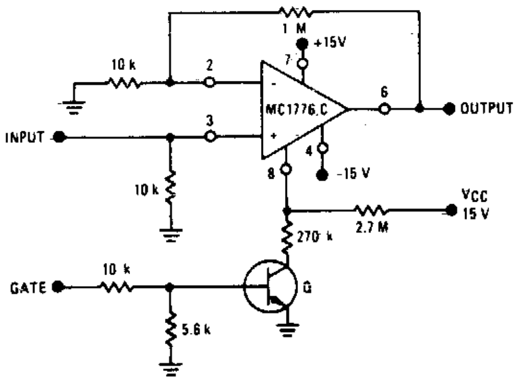
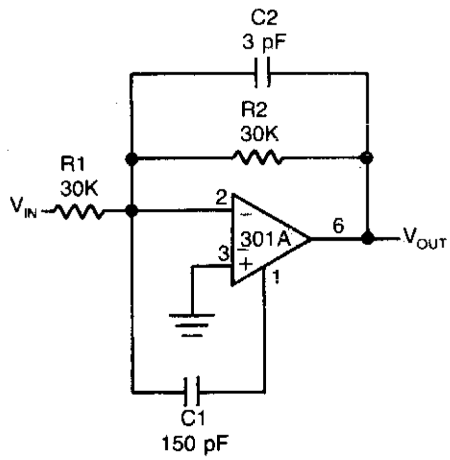


Fig. 3-25

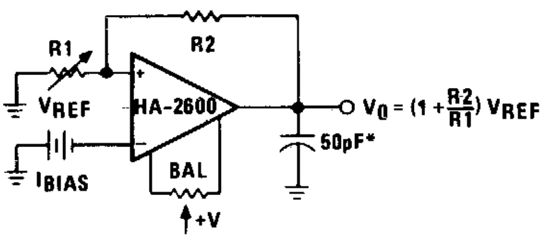
FAST SUMMING AMPLIFIER



Power Bandwidth: 250 kHz
Small Signal Bandwidth: 3.5 MHz
Slew Rate: 10V/μs

Fig. 3-27

REFERENCE VOLTAGE AMPLIFIER



FEATURES
1 MINIMUM BIAS CURRENT IN REFERENCE CELL
2 SHORT CIRCUIT PROTECTION

IC = HA-OP07

Fig. 3-26

ADJUSTMENT-FREE PRECISION SUMMING AMPLIFIER

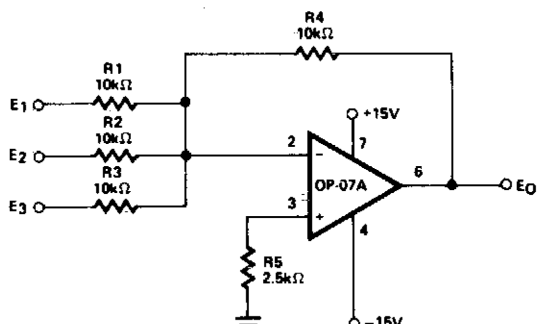
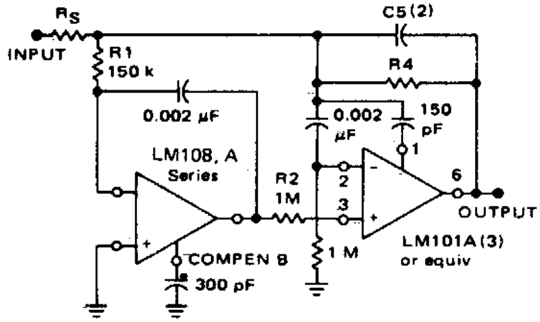


Fig. 3-28

Circuit Notes

This circuit produces continuous outputs that are a function of multiple input variables.

SUMMING AMPLIFIER WITH LOW INPUT CURRENT



(1) Power Bandwidth: 250 kHz
Small Signal Bandwidth: 3.5 MHz
Slew Rate: 10 V/µs

(2) $C5 = \frac{6 \times 10^{-8}}{R1}$

(3) In addition to increasing speed, the LM101A raises high and low frequency gain, increases output drive capability and eliminates thermal feedback.

Fig. 3-29

× 100 OPERATIONAL AMPLIFIER USING L161

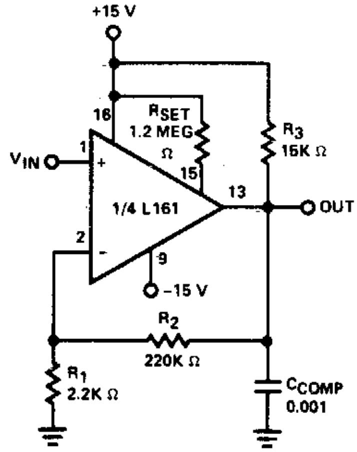


Fig. 3-31

Circuit Notes

Amplifier has gain-bandwidth product of 20 MHz with slew rate of 0.3V/µ sec.

× 10 OPERATIONAL AMPLIFIER USING L161

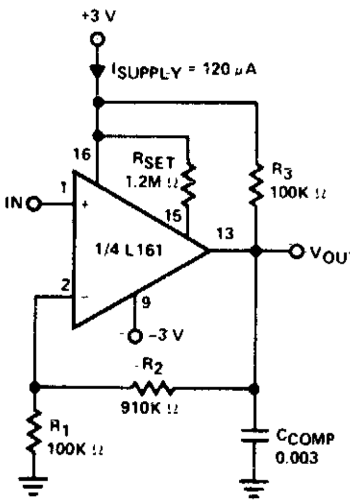
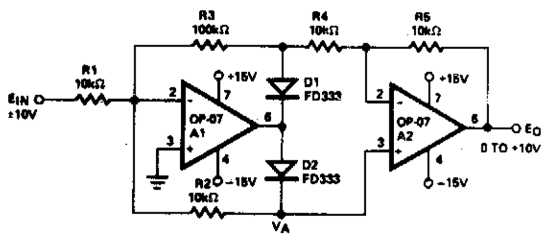


Fig. 3-30

Circuit Notes

Amplifier is 3 dB down at 100 kHz and has a slew rate of 0.02V/µ sec.

PRECISION ABSOLUTE VALUE CIRCUIT



POSITIVE INPUT

1. $V_A = 0$, D2 OFF, D1 ON
2. $E_O = \left(\frac{-E_{IN} R2}{R1}\right) \cdot \left(\frac{-R5}{R4}\right) = E_{IN} \frac{R3\,R5}{R1\,R4}$
3. WITH $R1 = R3 = R4 = R5$: $E_O = E_{IN}$
4. V_{OS} ERROR INCLUDED: $E_O = E_{IN} + 2V_{OS2}$

NEGATIVE INPUT

1. D1 OFF, D2 ON
2. $\frac{-E_{IN}}{R1} = \frac{V_A}{R2} + \frac{V_A}{R3 + R4}$
3. $E_O = V_A \left(1 + \frac{R5}{R3 + R4}\right)$
4. WITH $R3 = R4 = R5$: $E_O = 1.5V_A$
5. $E_O = -\frac{(R2)\,(R3 + R4)\,(1.5)\,E_{IN}}{R1\,(R2 + R3 + R4)}$
6. WITH $R1 = R2 = R3 = R4$: $E_O = -E_{IN}$
7. V_{OS} ERROR INCLUDED: $E_O = -E_{IN} + 1.5V_{OS2} - 0.5V_{OS1}$
8. FOR BOTH INPUTS: $E_O = +\ E_{IN}$

Fig. 3-32

ULTRA-LOW-LEAKAGE PREAMP

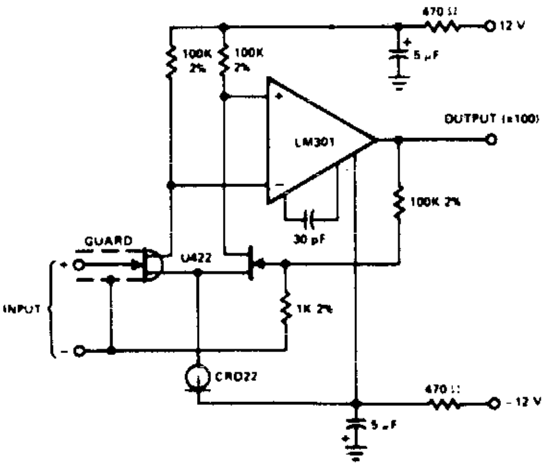
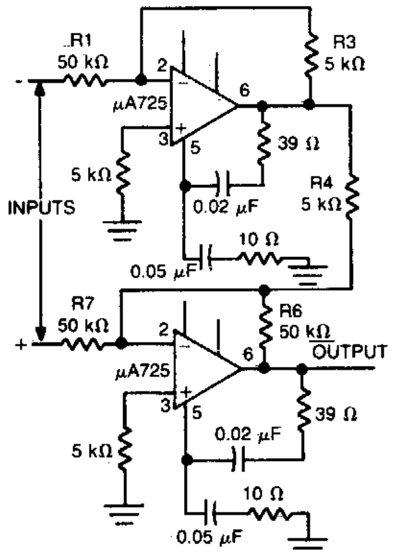


Fig. 3-33

Circuit Notes

Input leakage—2 pA at 75 °C.

±100 V COMMON MODE RANGE DIFFERENTIAL AMPLIFIER



Pin numbers are shown for metal package only.

Fig. 3-35

DC TO VIDEO LOG AMPLIFIER

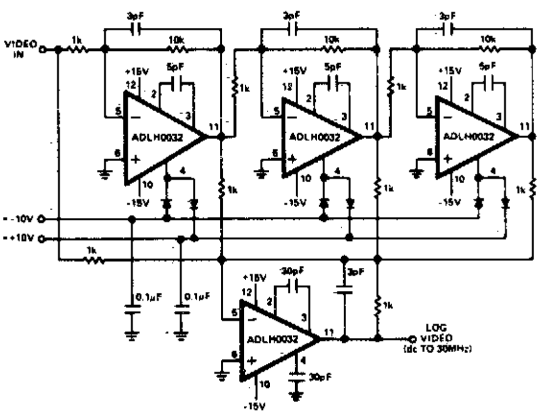
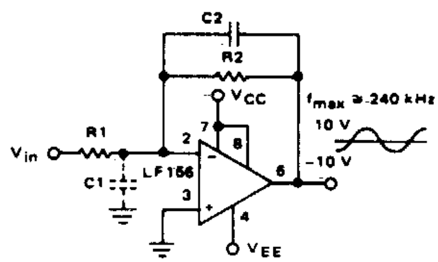


Fig. 3-34

WIDE BANDWIDTH, LOW NOISE, LOW DRIFT AMPLIFIER

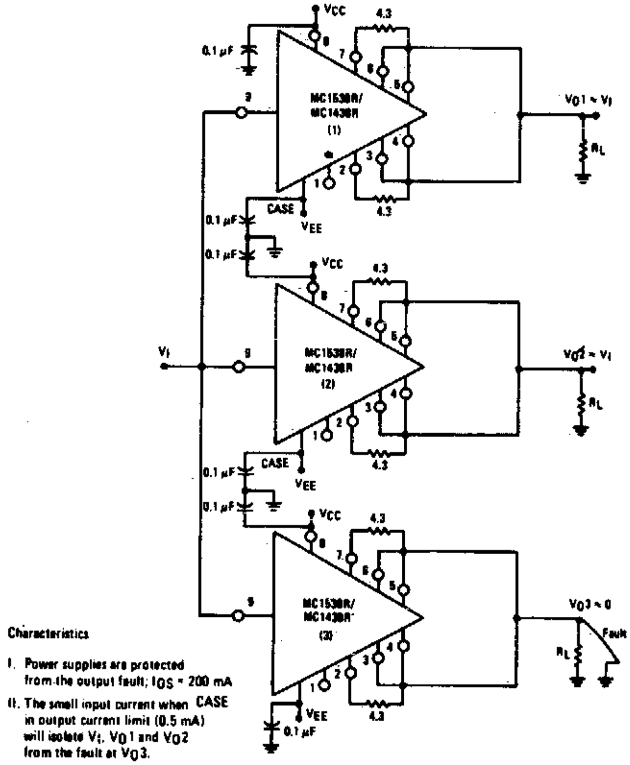


- Power BW: $f_{max} = \frac{S_r}{2\pi V_p} \cong 240$ kHz
- Parasitic input capacitance (C1 ≅ 3 pF for LF155, LF156, and LF157 plus any additional layout capacitance) interacts with feedback elements and creates undesirable high frequency pole. To compensate add C2 such that: R2C2 ≅ R1C1.

Fig. 3-36

SIGNAL DISTRIBUTION AMPLIFIER

Fig. 3-37



AUDIO DISTRIBUTION AMPLIFIER

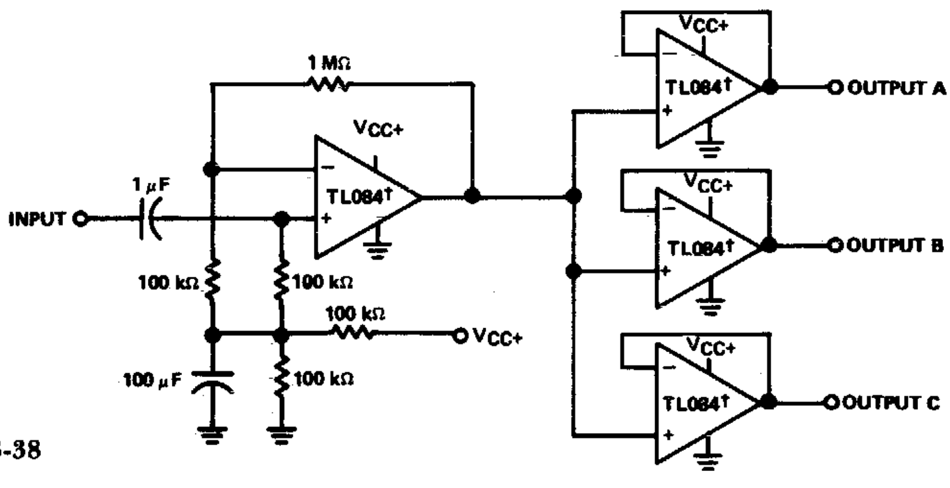


Fig. 3-38

HIGH INPUT IMPEDANCE, HIGH OUTPUT CURRENT VOLTAGE FOLLOWER

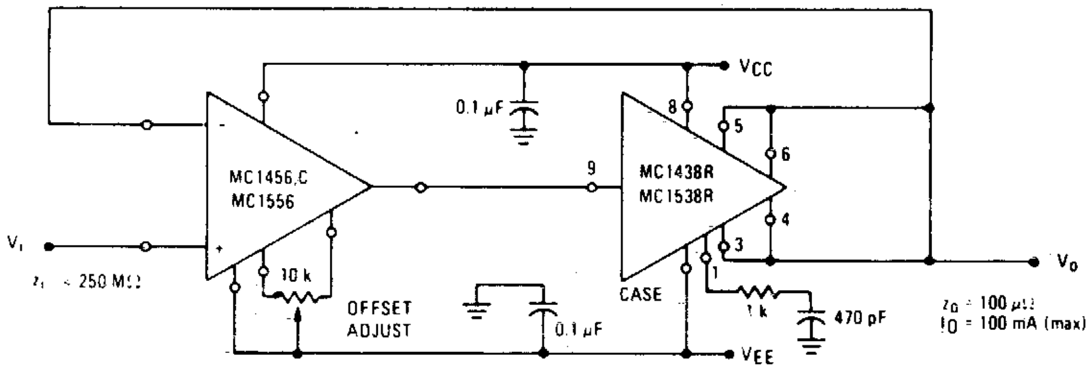
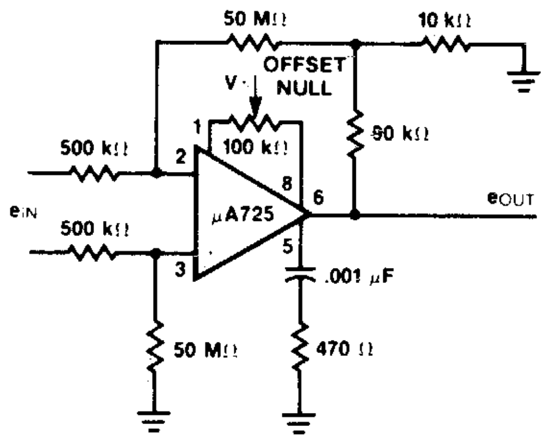


Fig. 3-39

PRECISION AMPLIFIER



Pin numbers are shown for metal package only.

Characteristics

$A_V = 1000 = 60$ dB
DC Gain Error = 0.05%
Bandwidth = 1 kHz for −0.05% error
Diff. Input Res. = 1 MΩ
Typical amplifying capability
$e_{IN} = 10\ \mu V$ on $V_{CMI} = 1.0$ V
Caution: Minimize Stray Capacitance
$A_{VCL} = 1000$

Fig. 3-40

PREAMPLIFIER AND HIGH-TO-LOW IMPEDANCE CONVERTER

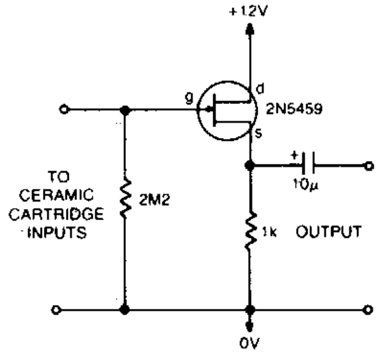


Fig. 3-41

Circuit Notes

Thiscircuit matches the very high impedance of ceramic cartridges, unity gain, and low impedance output. By "loading" the cartridge with a 2M2 input resistance, the cartridge characteristics are such as to quite closely compensate for the RIAA recording curve. The output from this preamp may be fed to a level pot for mixing.

NONINVERTING AMPLIFIER

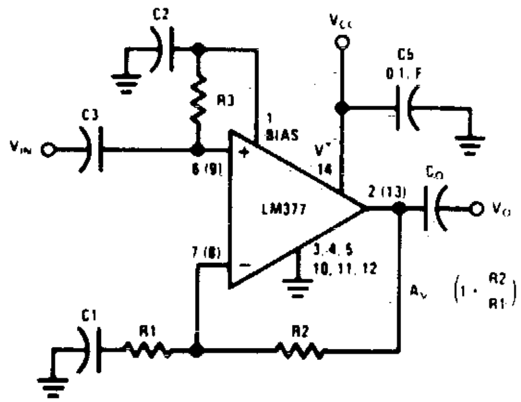


Fig. 3-42

HIGH IMPEDANCE, HIGH GAIN, HIGH FREQUENCY INVERTING AMP

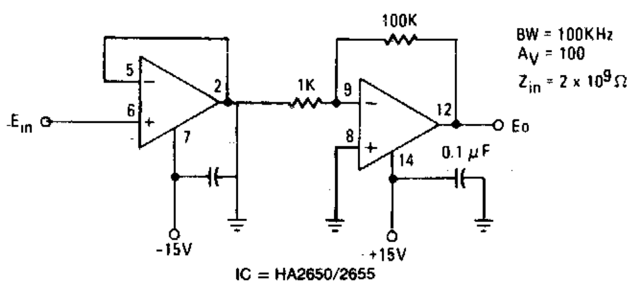


Fig. 3-43

LOG-RATIO AMPLIFIER

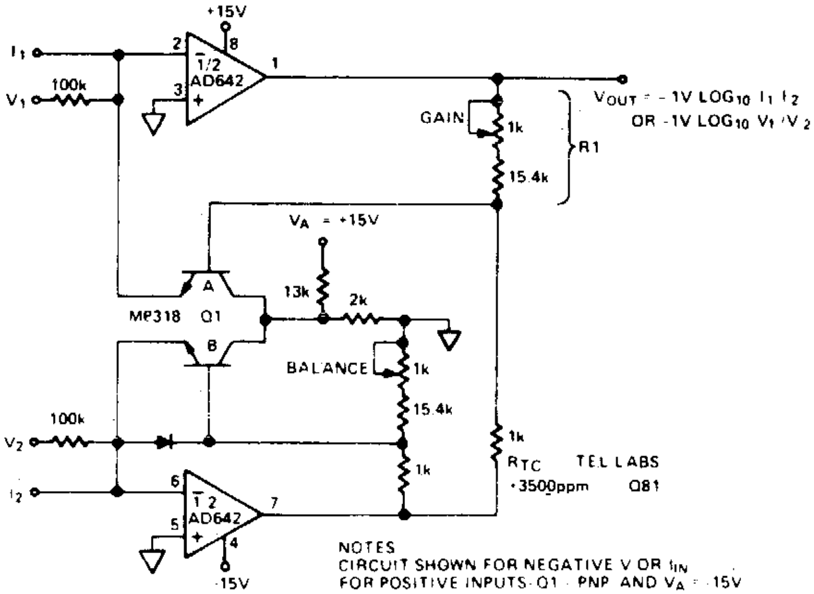


Fig. 3-44

INVERTING AMPLIFIER

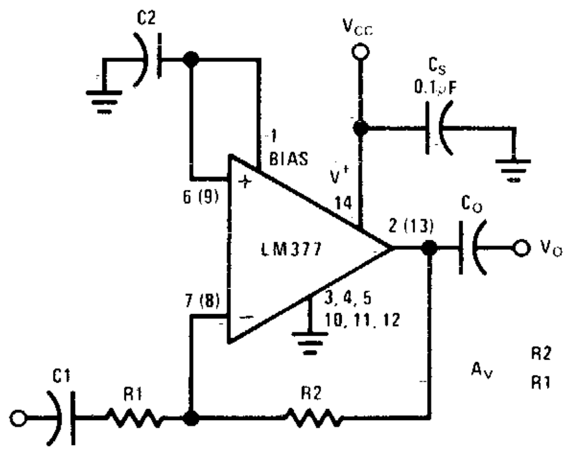


Fig. 3-45

4

Analog-to-Digital Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-BIT A/D CONVERTER

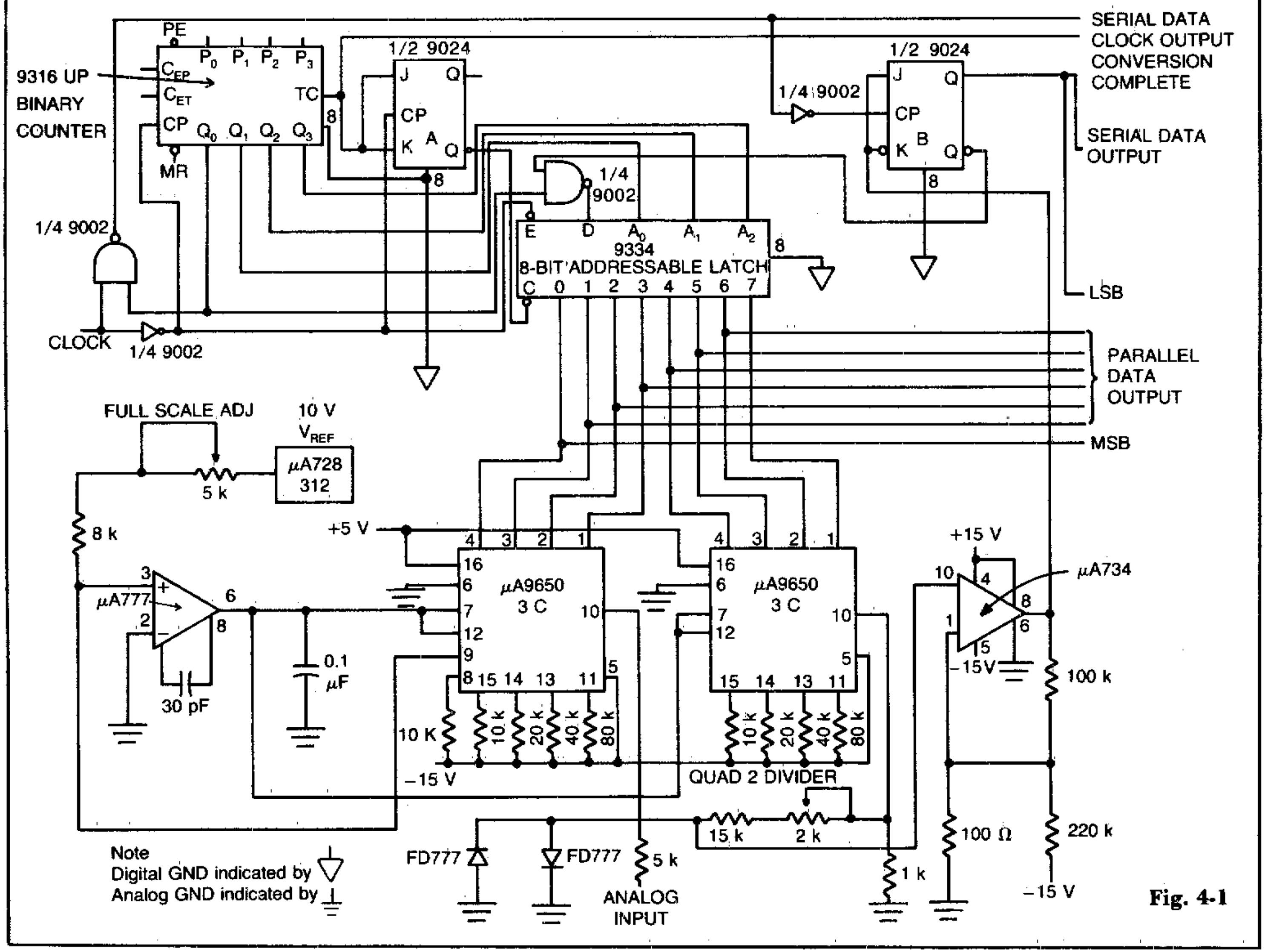


Fig. 4-1

SUCCESSIVE APPROXIMATION A/D CONVERTER

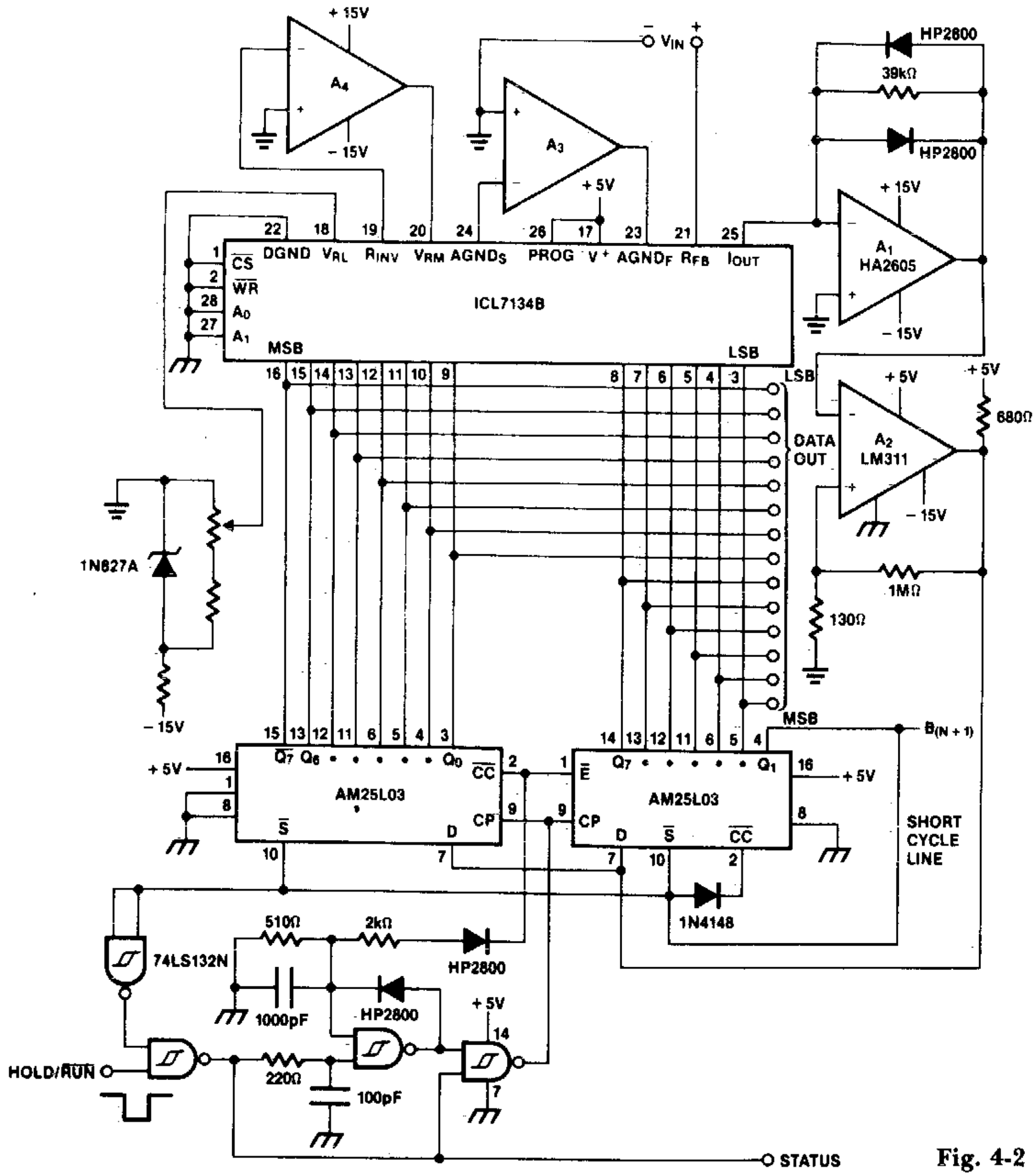


Fig. 4-2

Circuit Notes

A bipolar input, high speed A/D converter uses two AM25L03s to form a 14-bit successive approximation register. The comparator is a two-stage circuit with an HA2605 front-end amplifier used to reduce settling time problems at the summing node. Careful offset-nulling of this amplifier is needed.

8-BIT A/D CONVERTER

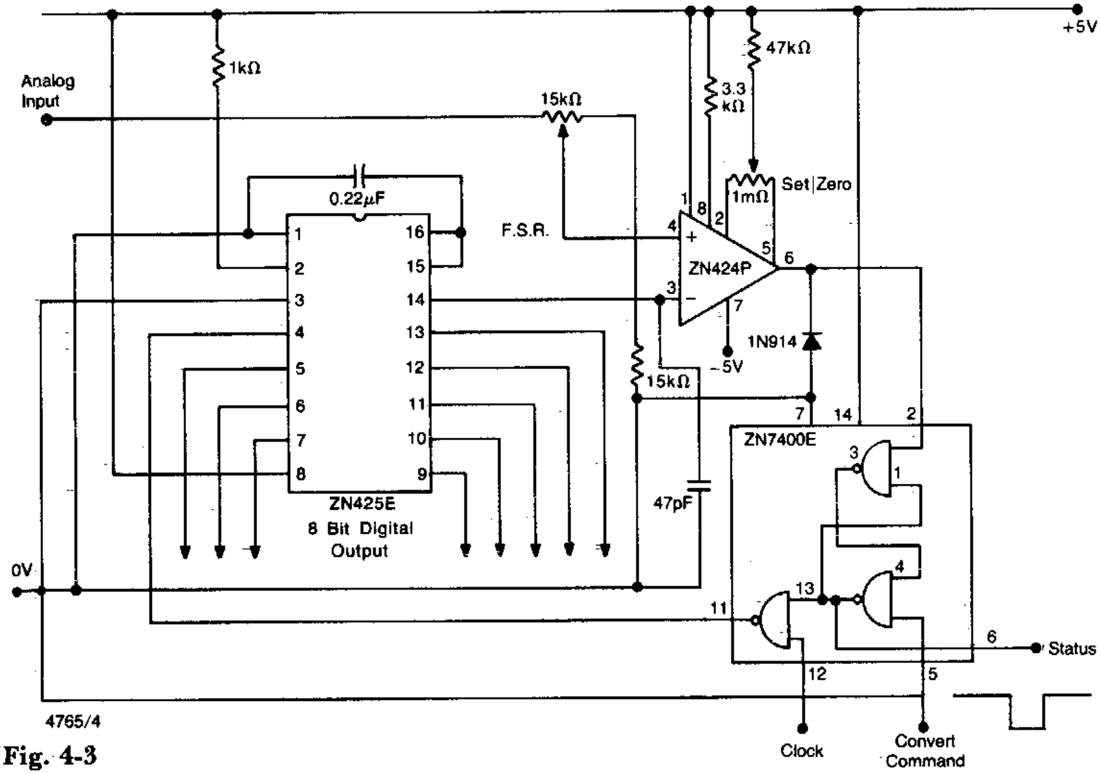


Fig. 4-3

8-BIT TRACKING A/D CONVERTER

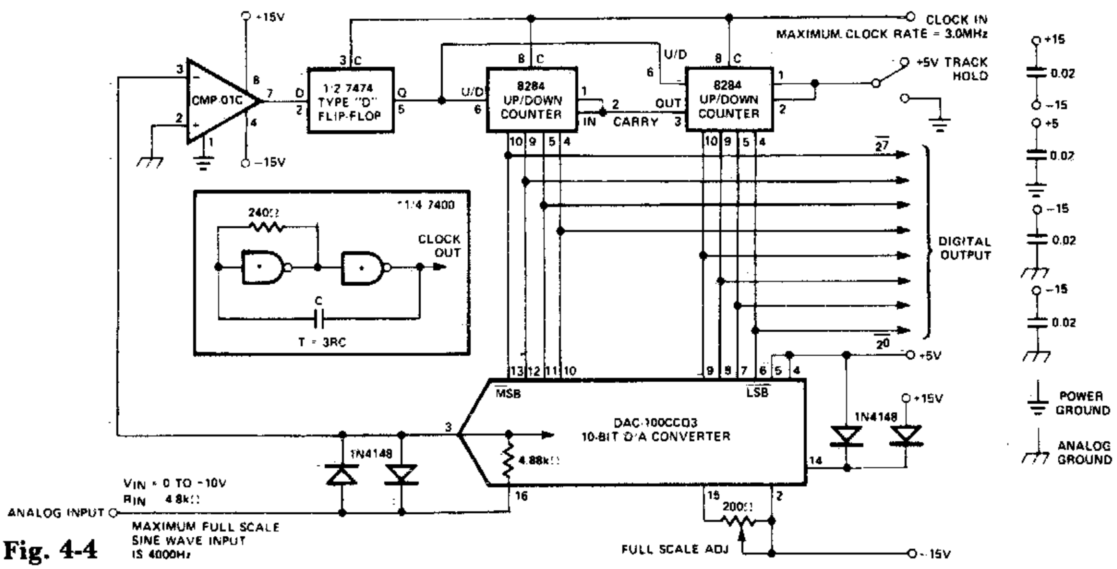


Fig. 4-4

8-BIT SUCCESSIVE APPROXIMATION A/D CONVERTER

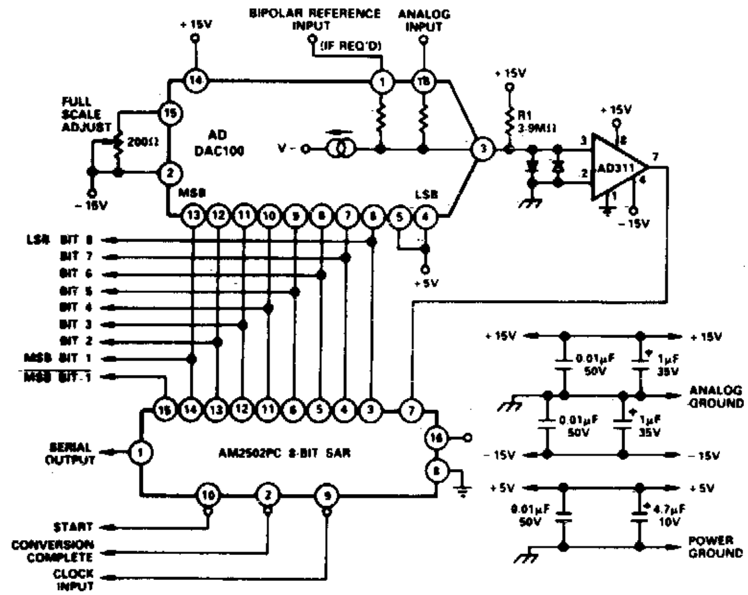


Fig. 4-5

FOUR CHANNEL DIGITALLY MULTIPLEXED RAMP A/D CONVERTER

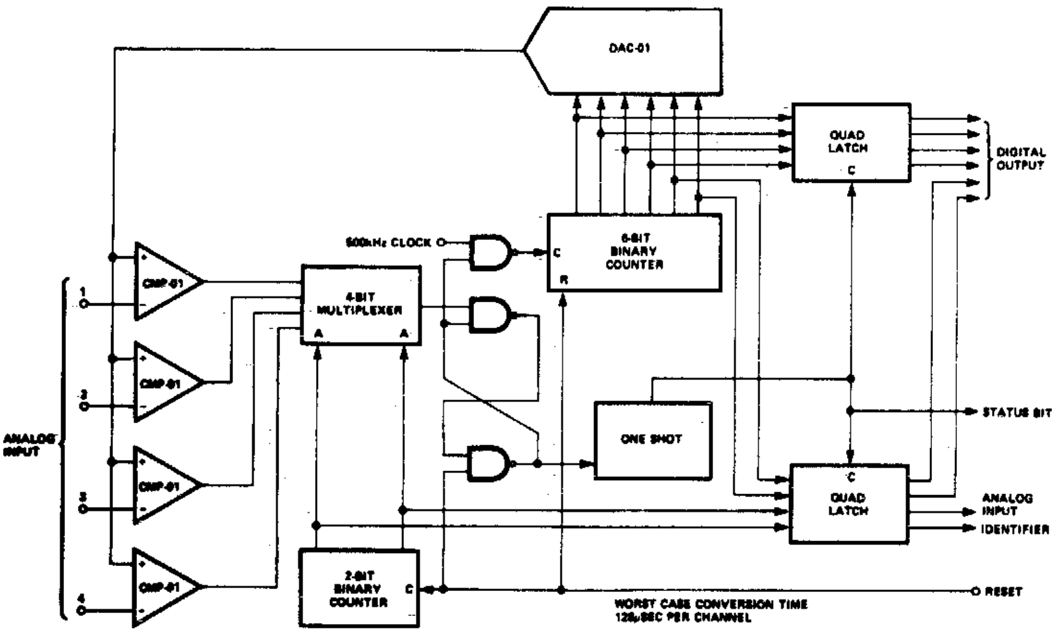


Fig. 4-6

THREE-DECADE LOGARITHMIC A/D CONVERTER

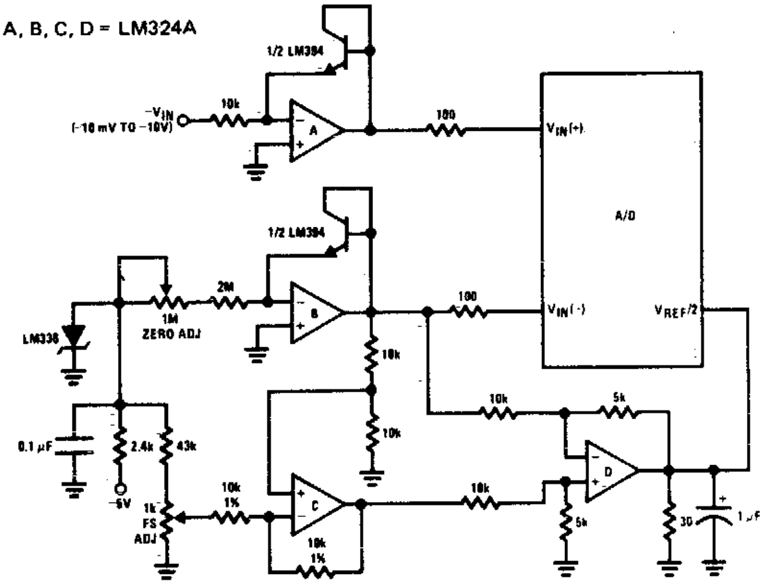


Fig. 4-7

TRACKING (SERVO TYPE) A/D CONVERTER

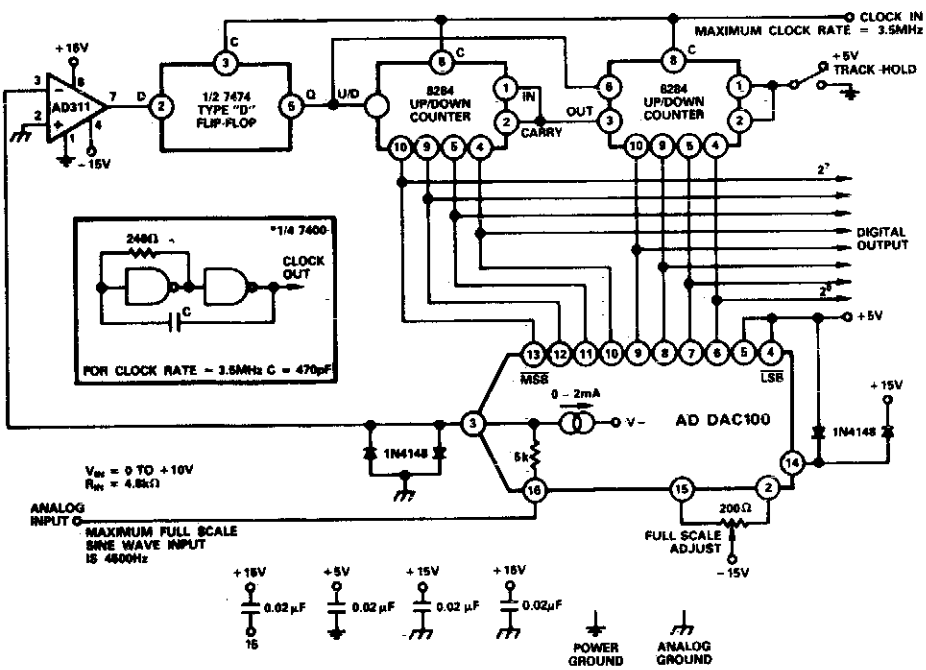


Fig. 4-8

3½ DIGIT A/D CONVERTER WITH LCD DISPLAY

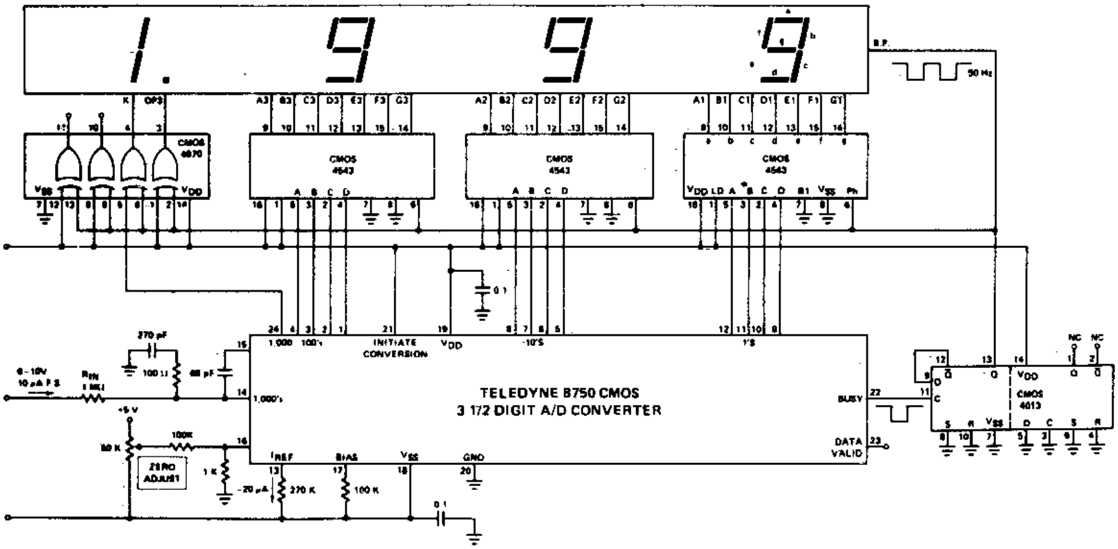
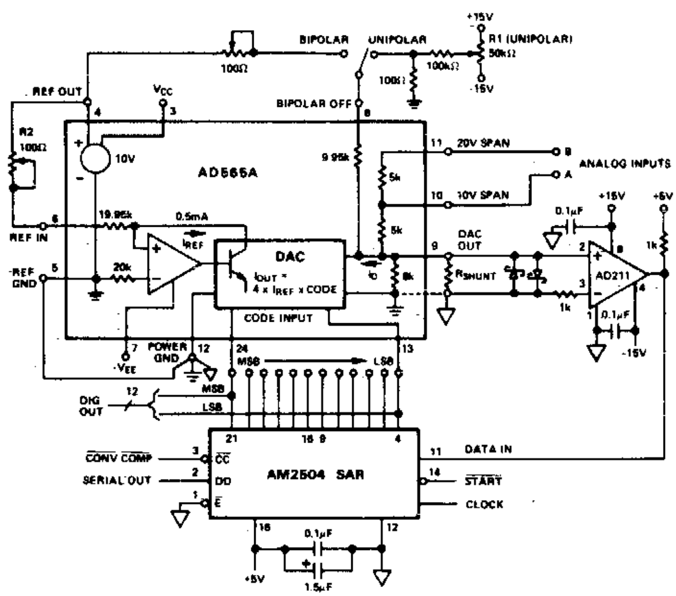


Fig. 4-9

FAST PRECISION A/D CONVERTER

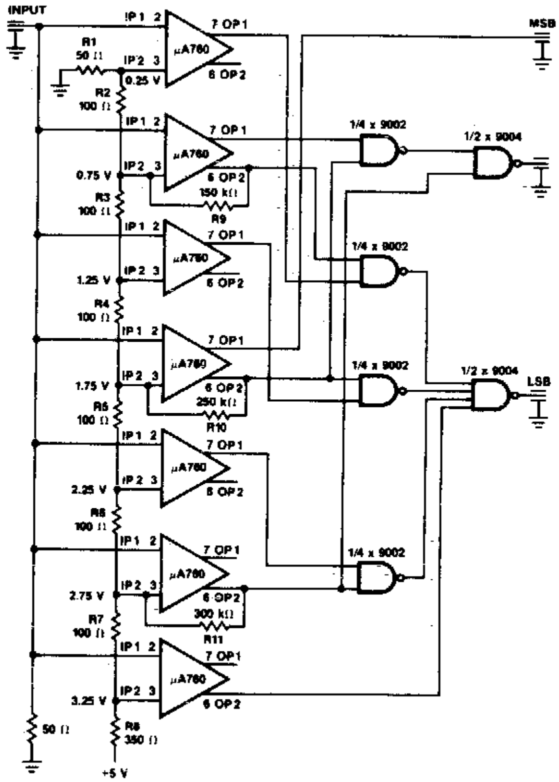


INPUT RANGES

Unipolar	Bipolar	Connect	Equiv DAC Z_{OUT}
0 TO +10	±5	INPUT TO A	2.35kΩ
0 TO +5	±2.5	INPUT TO A	1.90kΩ
0 TO +20	±10	INPUT TO B B TO DAC OUT	3.08kΩ

Fig. 4-10

HIGH SPEED 3-BIT A/D CONVERTER



Input voltage range: 3.5 V
Typical conversion speed: 30 ns

Fig. 4-11

THREE IC LOW COST A/D CONVERTER

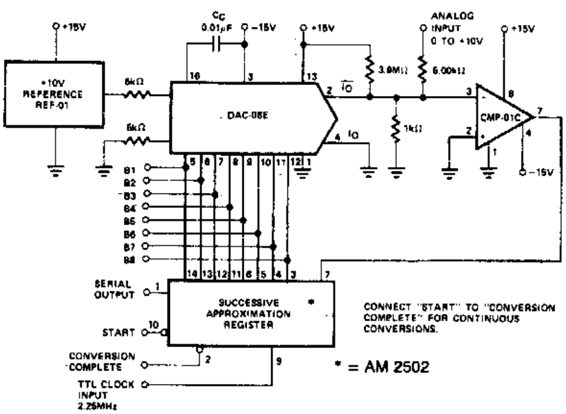


Fig. 4-12

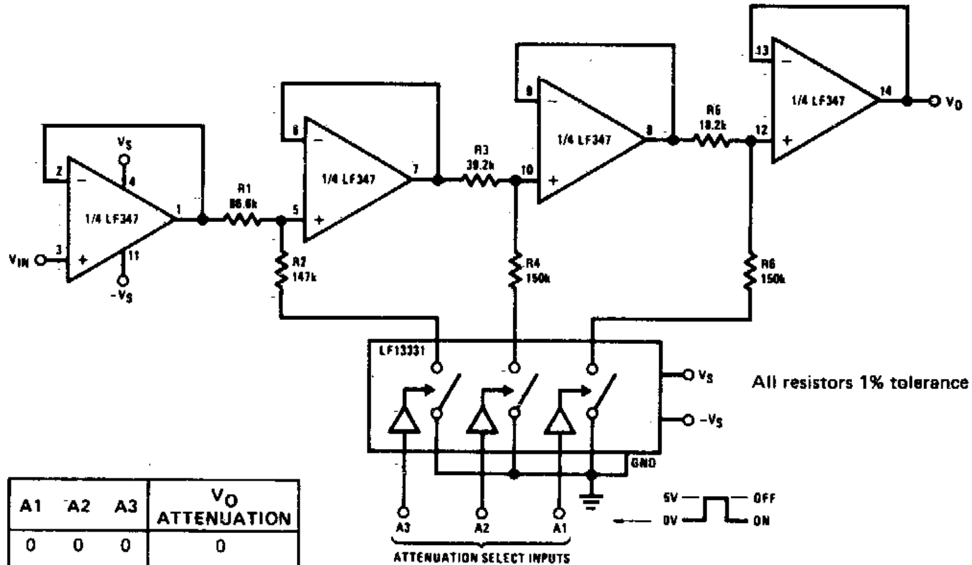
5

Attenuators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Digitally Selectable Precision Attenuator
Variable Attenuator
Digitally Controlled Amplifier/Attenuator
Programmable Attenuator (1 to 0.0001)

DIGITALLY SELECTABLE PRECISION ATTENUATOR



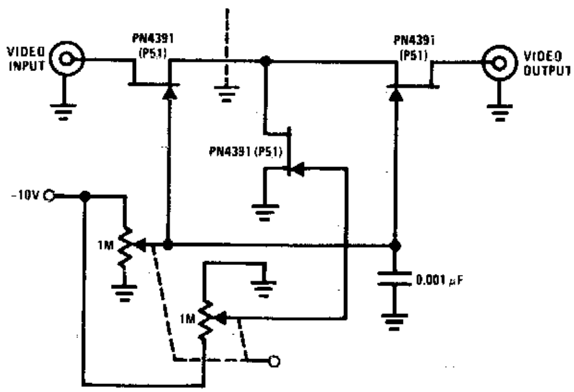
All resistors 1% tolerance

A1	A2	A3	V_O ATTENUATION
0	0	0	0
0	0	1	−1 dB
0	1	0	−2 dB
0	1	1	−3 dB
1	0	0	−4 dB
1	0	1	−5 dB
1	1	0	−6 dB
1	1	1	−7 dB

- Accuracy of better than 0.4% with standard 1% value resistors
- No offset adjustment necessary
- Expandable to any number of stages
- Very high input impedance

Fig. 5-1

VARIABLE ATTENUATOR

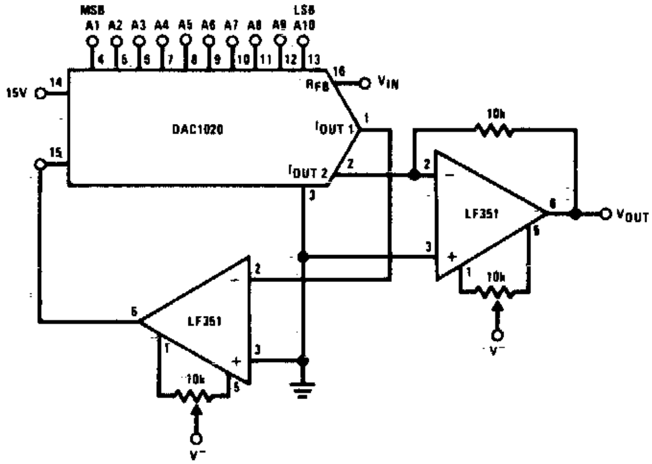


Circuit Notes

The PN4391 provides a low $R_{ds(on)}$ (less than 30 ohms). The tee attenuator provides for optimum dynamic linear range for attenuation and if complete turn-off is desired, attenuation of greater than 100 dB can be obtained at 10 MHz providing proper rf construction techniques are employed.

Fig. 5-2

DIGITALLY CONTROLLED AMPLIFIER/ATTENUATOR



$$V_{OUT} = V_{REF}\left[\frac{\frac{\overline{A1}}{2} + \frac{\overline{A2}}{4} + \ldots + \frac{A10}{1024}}{\frac{A1}{2} + \frac{A2}{4} + \ldots + \frac{\overline{A10}}{1024}}\right] \text{ or } V_{OUT} = V_{REF}\left(\frac{1023 - N}{N}\right)$$

where: $0 \le N \le 1023$
$N = 0$ for A_N = all zeros
$N = 1$ for $A10 = 1$, $A1-A9 = 0$
.
.
.
$N = 1023$ for A_N = all 1's

Fig. 5-3

PROGRAMMABLE ATTENUATOR (1 TO 0.0001)

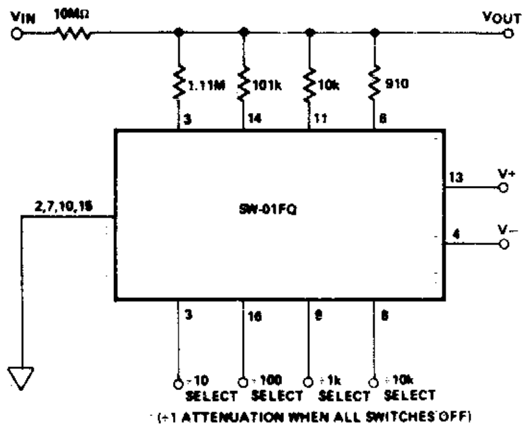


Fig. 5-4

6

Audio Mixers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FOUR-INPUT STEREO MIXER

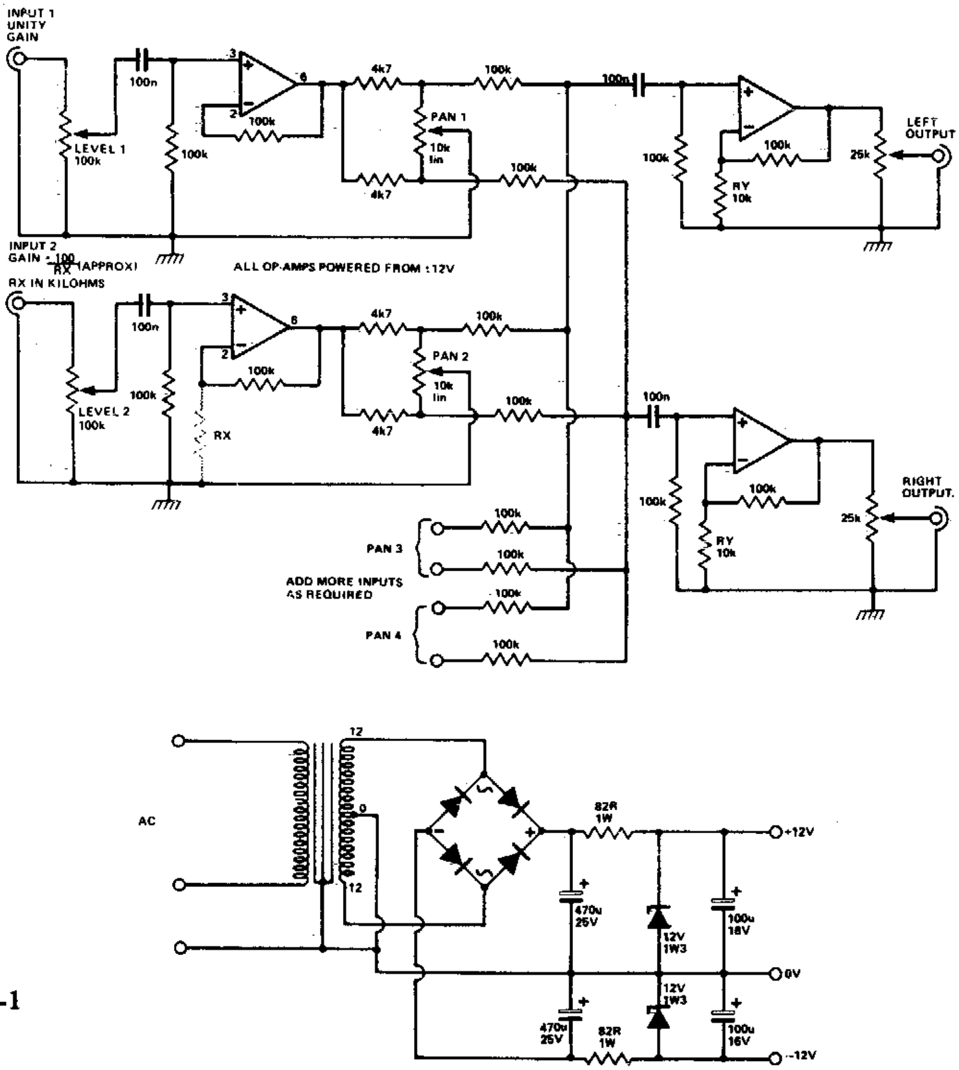
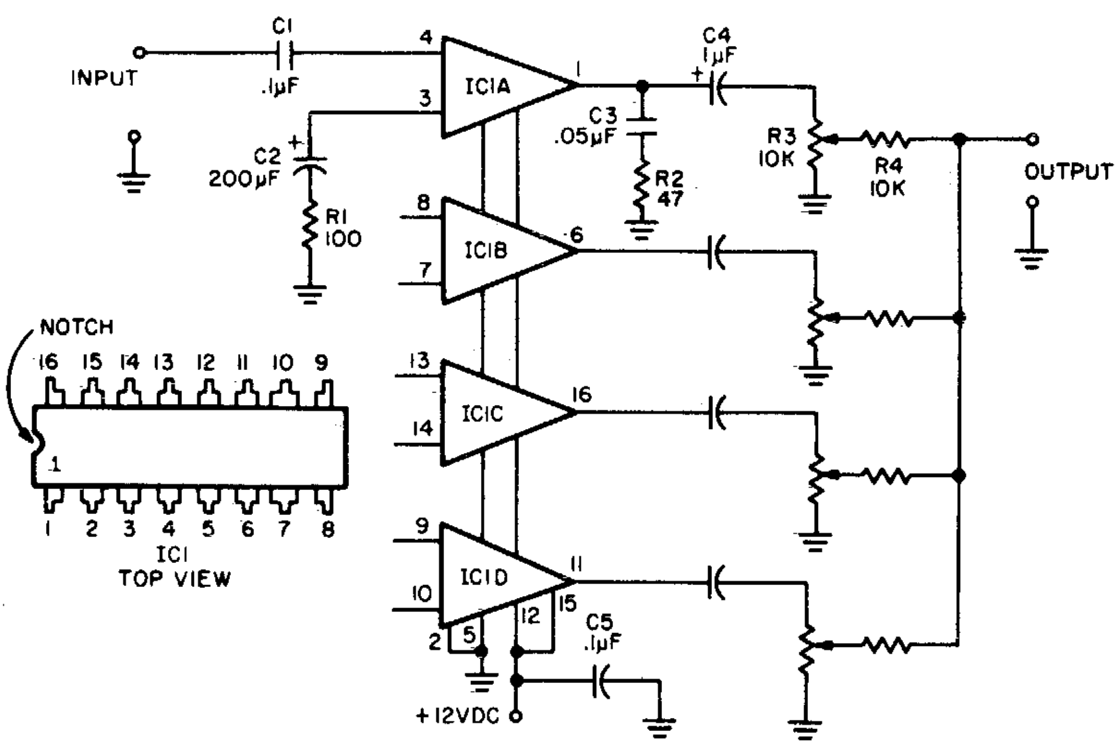


Fig. 6-1

Circuit Notes

Four (or more) inputs can be mixed and produce stereo output. Gain of each stage can be boosted by adding RX, but it should be kept below 50 (RX above 2.2 K) to avoid poor frequency, response. If more than four stages are used, decrease RX to 6.8 K for six inputs, or 4.7 K for eight inputs. The op amps are 741 or other lower noise types. The power supply circuit is also given.

HIGH-LEVEL FOUR-CHANNEL MIXER



PARTS LIST FOR HI-LEVEL MIXER

C1—0.1-uF, 3 VDC capacitor
C2—200-uF, 3 VDC capacitor
C3—0.05-uF, 75 VDC disc capacitor
C4—1-uF, 15 VDC capacitor
C5—0.1-uF, 15 VDC capacitor
IC1—RCA CA 3052
R1—100-ohms, ½-watt resistor
R2—47-ohms, ½-watt resistor
R3—Potentiometer, 10,000-ohms audio taper
R4—10,000-ohms, ½-watt resistor

Fig. 6-2

Circuit Notes

To provide good signal-to-noise ratio, this four channel mixer amplifier controls the signal levels after the amplifiers, and then mixes them to offer a combined output. The circuit works with any 50 ohm to 50 K dynamic microphone but not with crystal or ceramic mikes because the IC input impedance is low. Note that all four circuits are identical but that only one is shown complete.

TWO CHANNEL PANNING CIRCUIT

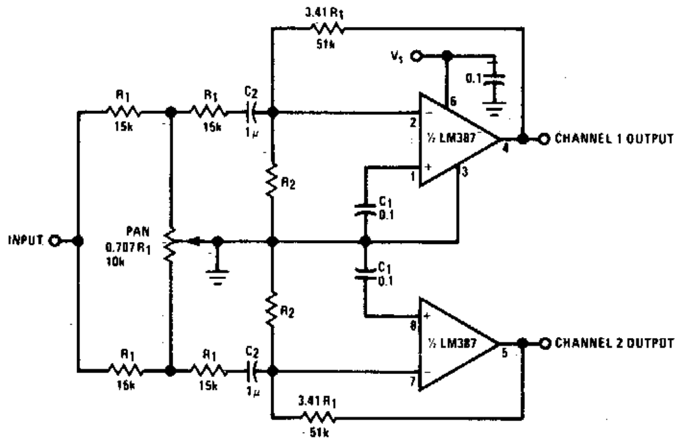


Fig. 6-3

Circuit Notes

This panning circuit (short for panoramic control circuit) provides the ability to move the apparent position of one microphone's input between two output channels. This effect is often required in recording studio mixing consoles. Panning is how recording engineers manage to pick up your favorite pianist and "float" the sound over to the other side of the stage and back again.

CMOS MIXER

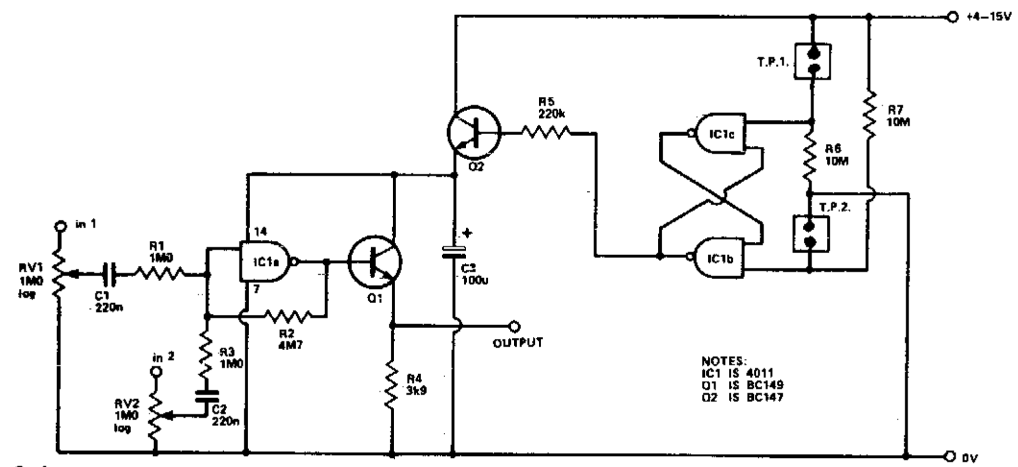


Fig. 6-4

Circuit Notes

Four inputs can be mixed by duplicating the circuit to the left of C3 and using the fourth gate of IC1. Two gates are used in a touch-operated switching circuit that controls the voltage on the base of switching transistor Q2. Touching TP1 and TP2 alternately turns the circuit on and off.

MIXER PREAMPLIFIER WITH TONE CONTROL

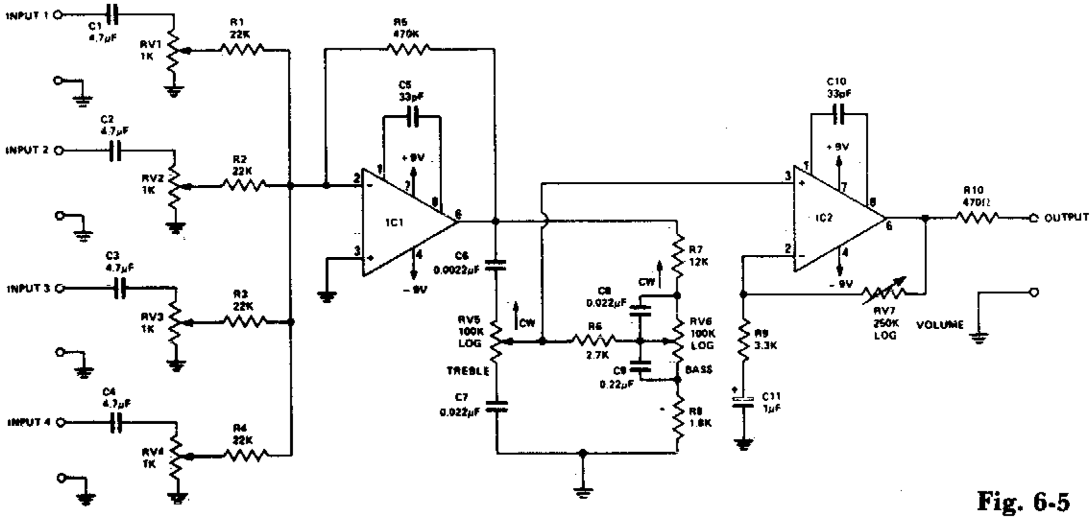


Fig. 6-5

Circuit Notes

General purpose preamplifier/mixer accepts up to four inputs, has a gain of 1600, and provides bass and treble controls that can be varied ± 10 dB at 100 Hz and 10 kHz respectively. IC1 and IC2 = LM301A.

PASSIVE MIXER

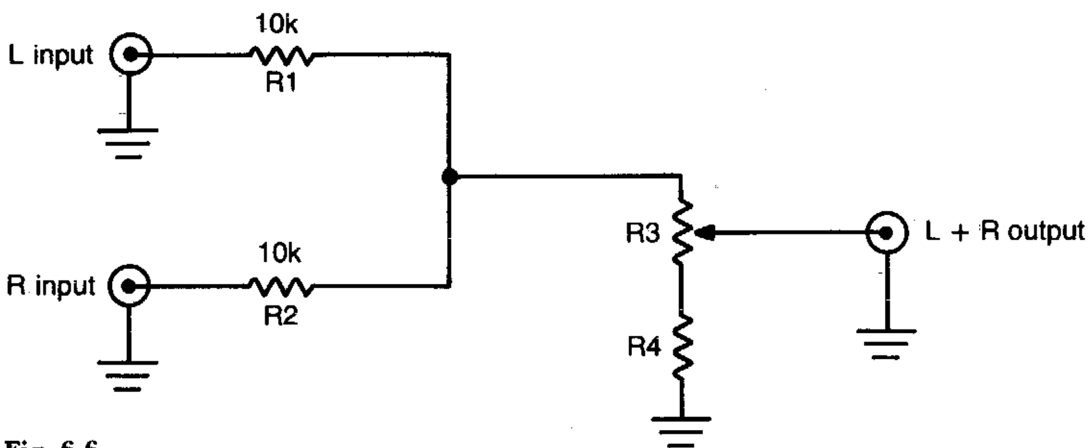


Fig. 6-6

Circuit Notes

This simple circuit can be used to combine stereo signals to produce a monaural output. R1 and R2 isolate both circuits and R3 controls the level of the combined output signal.

ONE TRANSISTOR AUDIO MIXER

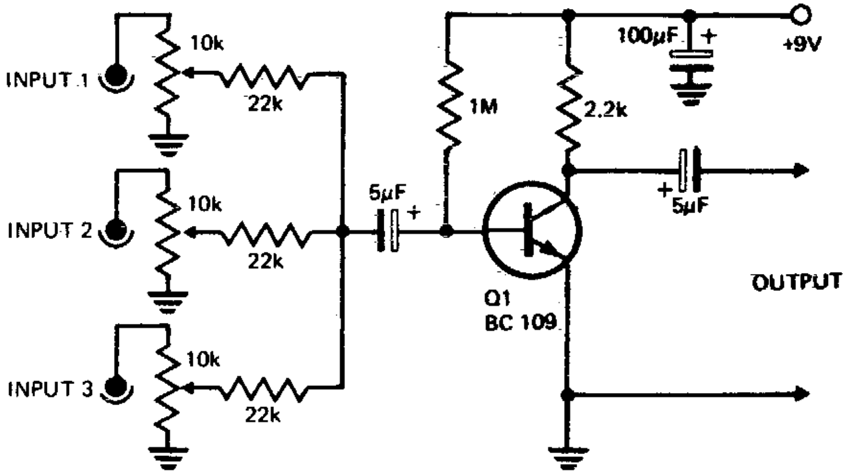


Fig. 6-7

Circuit Notes

Three or more inputs with individual level controls feed into the base of Q1 that provides a voltage gain of 20.

SILENT AUDIO SWITCHING/MIXING

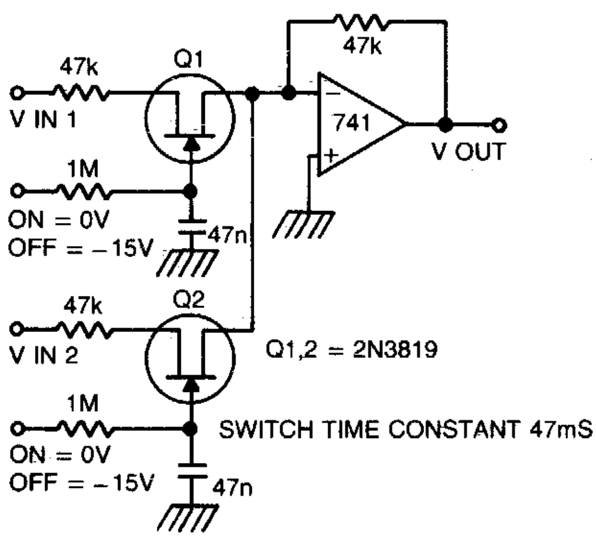


Fig. 6-8

Circuit Notes

Two or more signals can be switched and/or mixed without annoying clicks by using FETs and a low input-impedance op amp circuit.

HYBRID MIXER

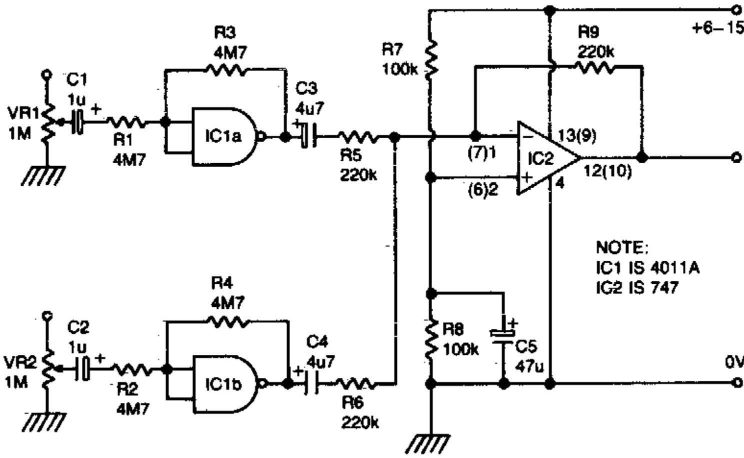


Fig. 6-9

Circuit Notes

IC1a and b are biased into the linear regions by R3 and R4. (IC1 must be 4011A). Outputs from gates are combined by op amp IC2, which provides low impedance output.

FOUR CHANNEL MIXER

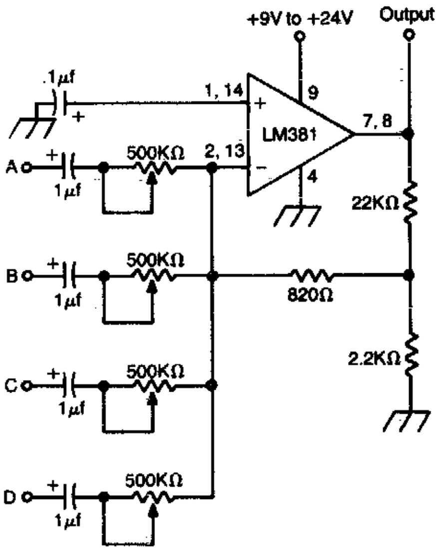


Fig. 6-10

Circuit Notes

High gain op amp combines up to four individually controlled input signals. The dc power source should be well filtered (battery is ideal), and the circuit should be well shielded to prevent hum pickup.

7

Audio Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

WIEN BRIDGE OSCILLATOR

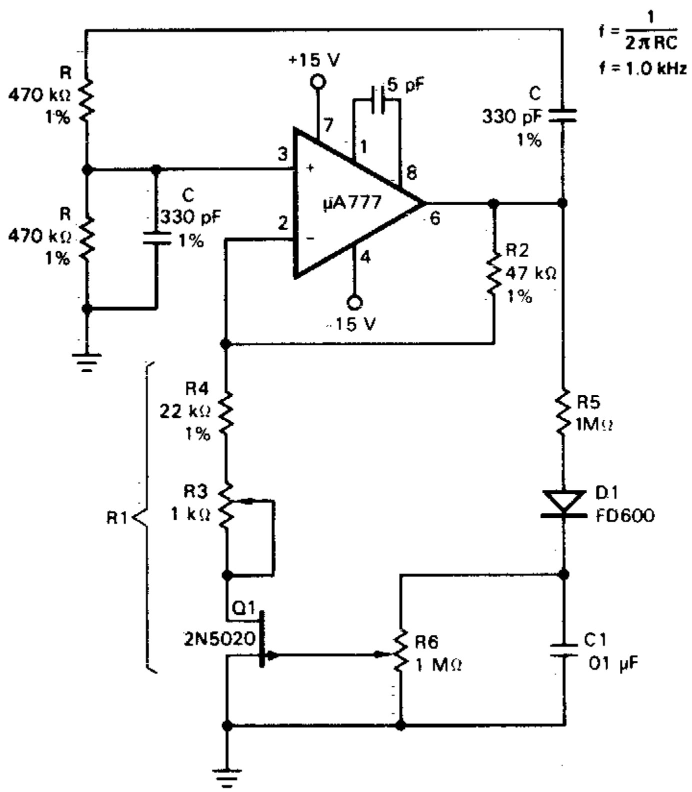


Fig. 7-1

Circuit Notes

Field effect transistor, Q1, operates in the linear resistive region to provide automatic gain control. Because the attenuation of the RC network is one-third at the zero phase-shift oscillation frequency, the amplifier gain determined by resistor R2 and equivalent resistor R1 must be just equal to three to make up the unity gain positive feedback requirement needed for stable oscillation. Resistors R3 and R4 are set to approximately 1000 ohm less than the required R1 resistance. The FET dynamically provides the trimming resistance needed to make R1 one-half of the resistance of R2. The circuit composed of R5, D1, and C1 isolates, rectifies, and filters the output sine wave, converting it into a dc potential to control the gate of the FET. For the low drain-to-source voltages used, the FET provides a symmetrical linear resistance for a given gate-to-source voltage.

WIEN BRIDGE OSCILLATOR

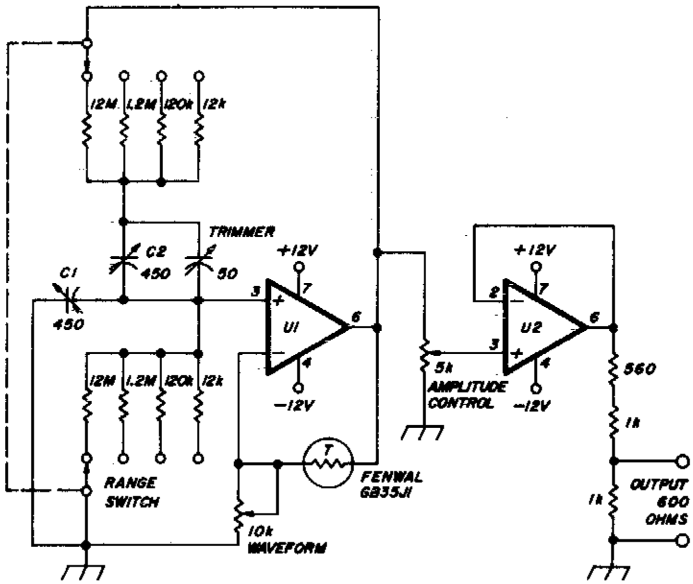


Fig. 7-2

Circuit Notes

Wien bridge sine-wave oscillator using two RCA CA3140 op amps covers 30 Hz to 100 kHz with less than 0.5 percent total harmonic distortion. The 10k pot is adjusted for the best waveform. Capacitor C1 and C2 are a two-gang, 450-pF variable with its frame isolated from ground. Maximum output into a 600-ohm load is about 1 volt rms.

WIEN BRIDGE OSCILLATOR

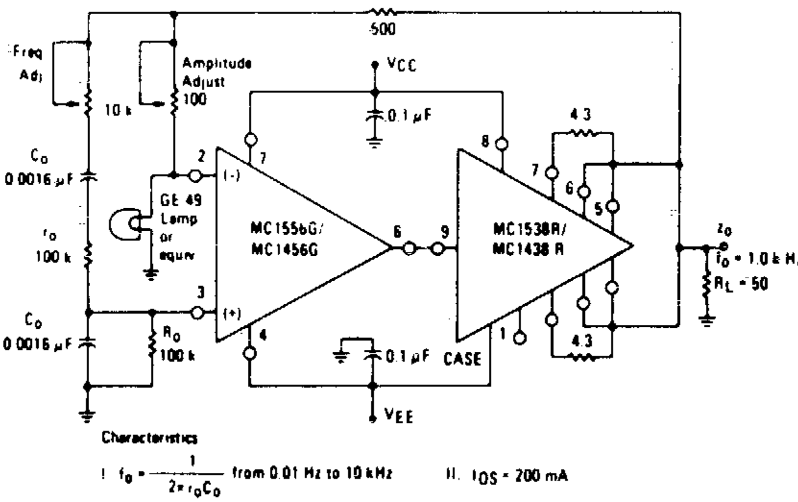
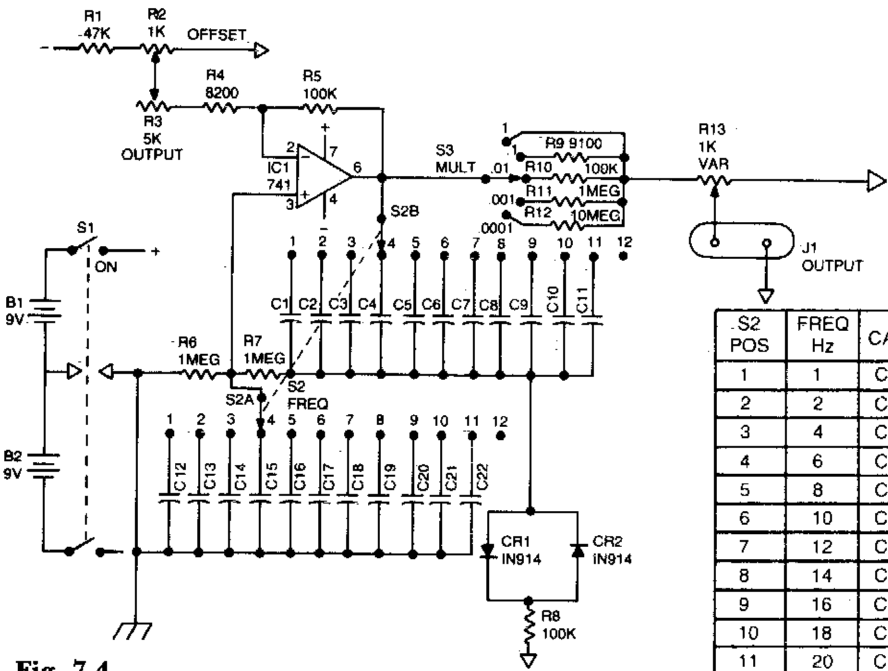


Fig. 7-3

VERY LOW FREQUENCY GENERATOR



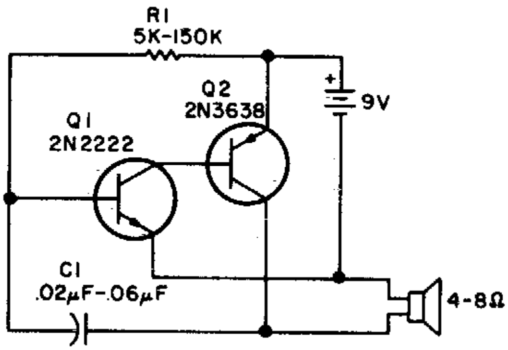
S2 POS	FREQ Hz	CAPACITOR	VALUE μF
1	1	C1=C12=	.5 + 1
2	2	C2=C13=	.15 + J5
3	4	C3=C14=	.15
4	6	C4=C15=	.1
5	8	C5=C16=	.066 + .0068
6	10	C6=C17=	.D5 + D1
7	12	C7=C 18=	.05
8	14	C8=C19=	.033 +.01
9	16	C9=C20=	0.33 +.0047
10	18	C10=C21=	.033
11	20	C11=C22=	.015+.015
12	–	–	–

Fig. 7-4

Circuit Notes

Wien bridge oscillator generates frequencies of 1 Hz and 2 to 20 Hz in 2 Hz steps. Maximum output amplitude is 3 volts rms of 8.5 volts peak-to-peak. A pot-and-switch attenuator allows the output level to be set with a fair degree of precision to any value within a range of 5 decades.

AUDIO OSCILLATOR



Circuit Notes

Almost any transistor will work. R1 and C1 will vary the tone.

Fig. 7-5

SINE WAVE OSCILLATOR

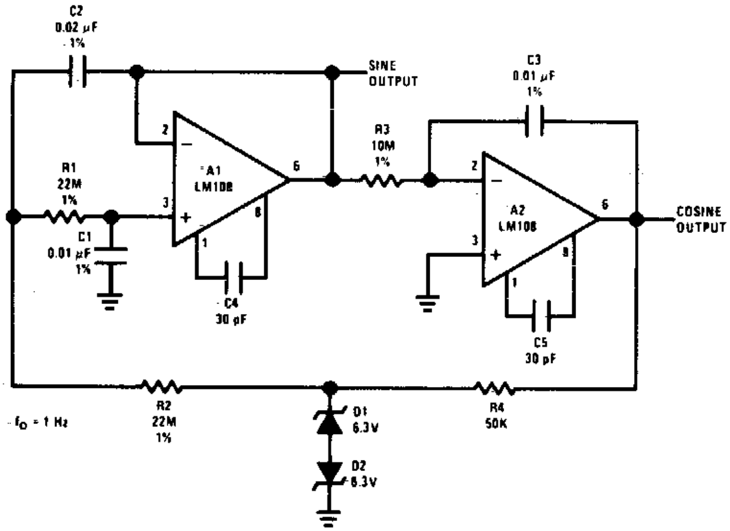
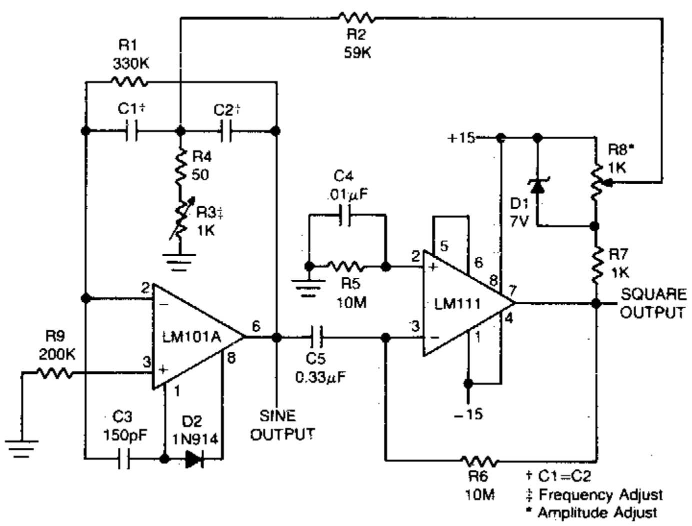


Fig. 7-6

Circuit Notes

The oscillator delivers a high-purity sinusoid with a stable frequency and amplitude.

EASILY TUNED SINE/SQUARE WAVE OSCILLATORS



$$F_o = \frac{1}{2\pi C_1 \sqrt{R_3 R_1}}$$

Fig. 7-7

Circuit Notes

This circuit will provide both a sine and square wave output for frequencies from below 20 Hz to above 20 kHz. The frequency of oscillation is easily tuned by varying a single resistor.

WIEN BRIDGE SINE WAVE OSCILLATOR

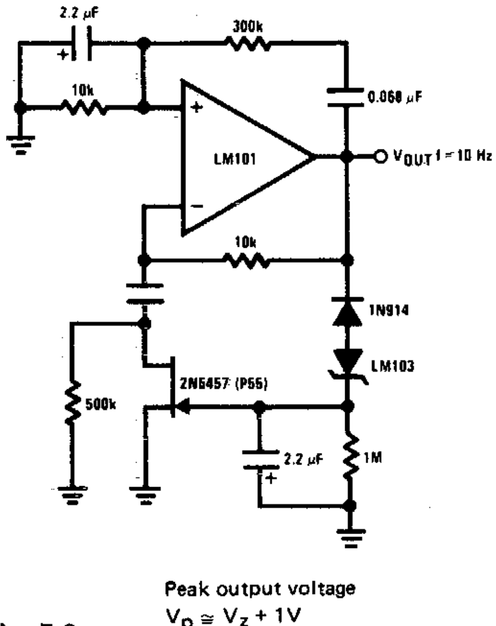


Fig. 7-8

Circuit Notes

Using the 2N5457 JFET as a voltage variable resistor in the amplifier feedback loop, produces a low distortion, constant amplitude sine wave getting the amplifier loop gain just right. The LM103 zener diode provides the voltage reference for the peak sine wave amplitude.

PHASE-SHIFT OSCILLATOR

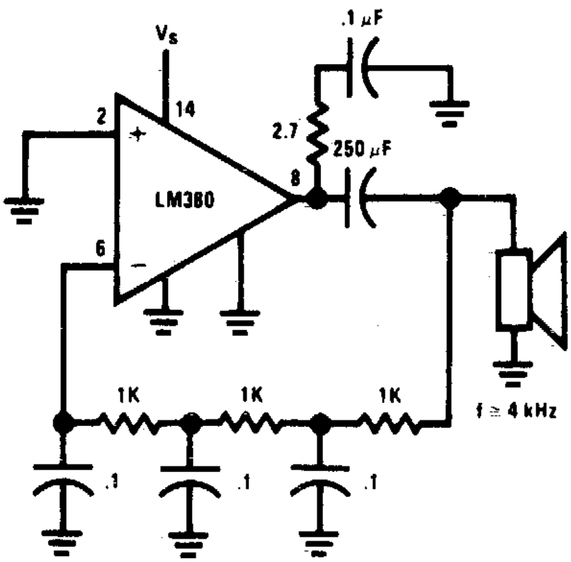


Fig. 7-9

Circuit Notes

Circuit uses a simple RC network to produce an exceptionally shrill tone from a miniature speaker. With the parts values shown, the circuit oscillates at a frequency of 3.6 kHz and drives a miniature 2½" speaker with ear-piercing volume. The output waveform is a square wave with a width of 150 μs, sloping rise and fall times, and a peak-to-peak amplitude of 4.2 volts (when powered by 9 volts). Current drain of the oscillator is 90 mA at 9 volts, and total power dissipation at this voltage is 0.81 watt, which is well below the 1.25 watts the 14-pin version will absorb (at room temperature) before shutting down.

TONE ENCODER

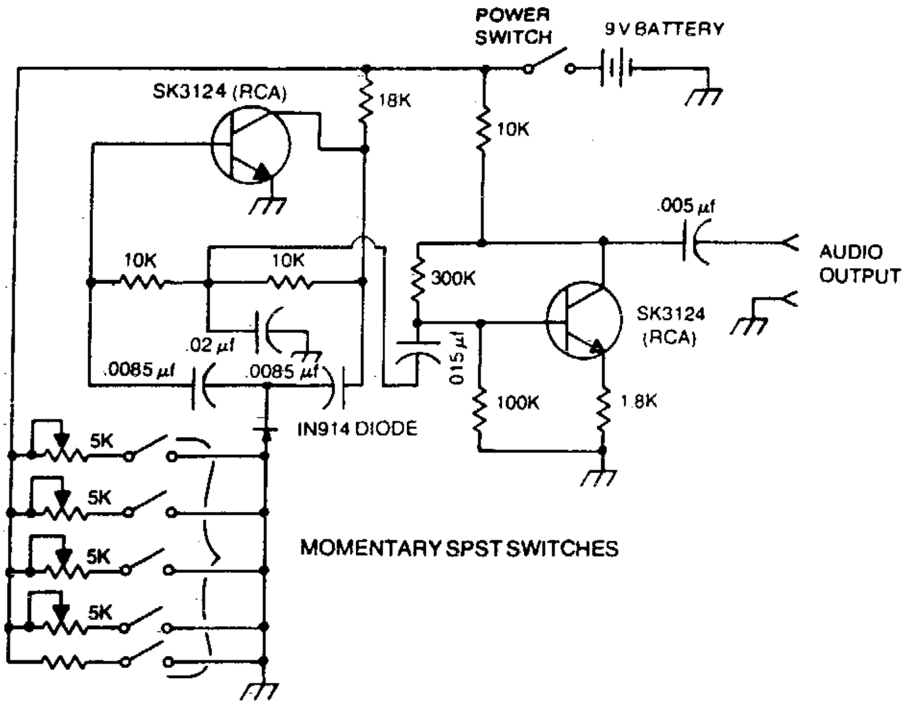
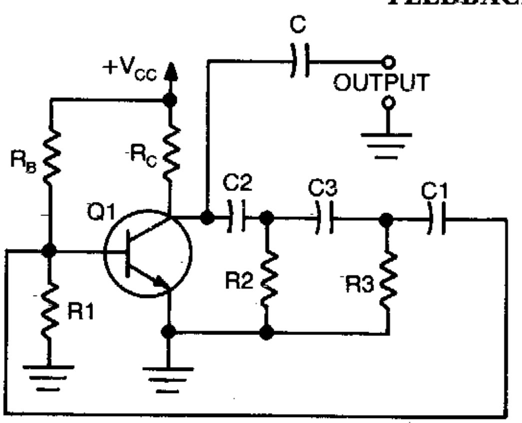


Fig. 7-10

Circuit Notes

A basic twin-T circuit uses resistors for accurately setting the frequency of the output tones, selected by pushbutton. Momentary switches produce a tone only when the button is depressed.

FEEDBACK OSCILLATOR



Circuit Notes

Circuit oscillates because the transistor shifts the phase of the signal 180° from the base to the collector. Each of the RC networks in the circuit is designed to shift the phase 60° at the frequency of oscillation for a total of 180°. The appropriate values of R and C for each network is found from $f = 1/2\sqrt{3}\pi RC$); that equation allows for the 60° phase shift required by the design.

Fig. 7-11

PHASE SHIFT OSCILLATOR

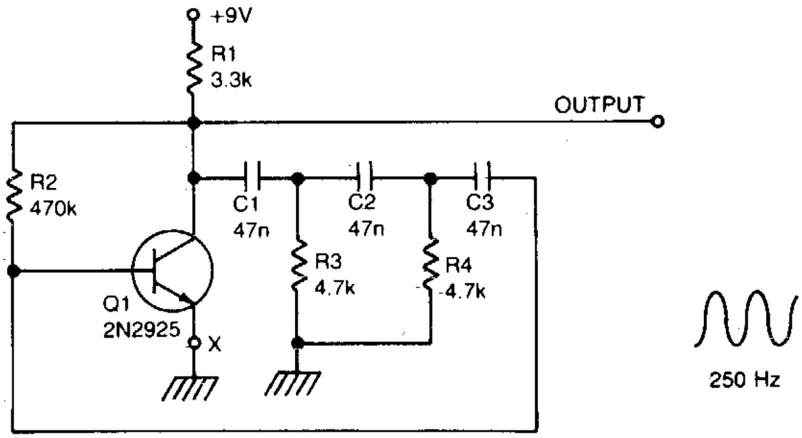


Fig. 7-12

Circuit Notes

A single transistor makes a simple phase shift oscillator. The output is a sine wave with distortion of about 10%. The sine wave purity can be increased by putting a variable resistor (25 ohms) in the emitter lead of Q1 (x). The resistor is adjusted so the circuit is only just oscillating, then the sine wave is relatively pure. Operating frequency may be varied by putting a 10 K variable resistor in series with R3, or by changing C1, C2, and C3. Making C1, 2, 3 equal to 100 nF will halve the operating frequency. Operating frequency can also be voltage controlled by a FET in series with R3, or optically controlled by an LDR in series with R3.

800 Hz OSCILLATOR

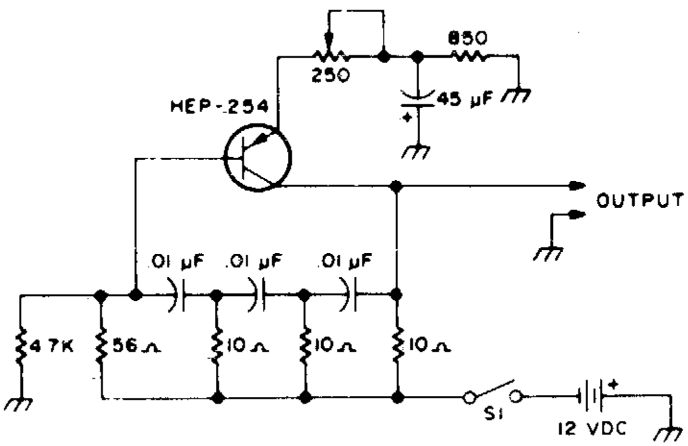


Fig. 7-13

Circuit Notes

The following transistors may be used: HEP-254, O.C-2, SK-3004, AT30H. To increase the frequency, decrease the value of the capacitors in the ladder network.

TUNABLE SINGLE COMPARATOR OSCILLATOR

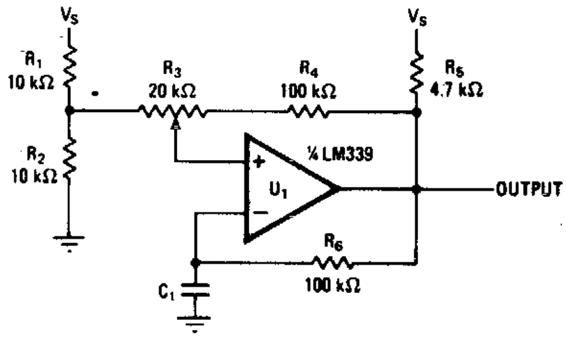


Fig. 7-14

Circuit Notes

Varying the amount of this comparator circuit's hysteresis makes it possible to vary output frequencies in the 740-Hz to 2.7-kHz range smoothly. The amount of hysteresis together with time constant R6C1 determines how much time it takes for C1 to charge or discharge to the new threshold after the output voltage switches.

WIDE RANGE OSCILLATOR (FREQUENCY RANGE OF 5000 TO 1)

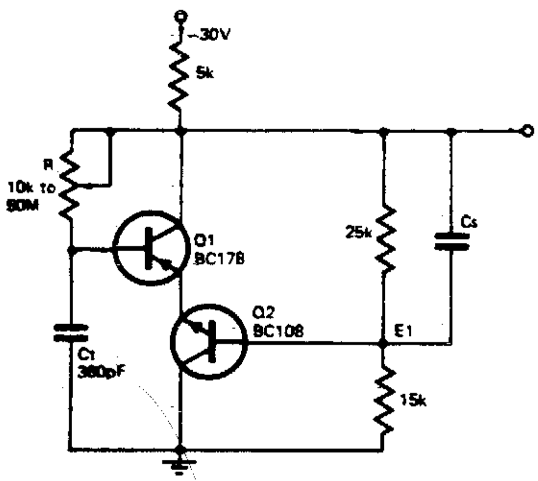


Fig. 7-15

Circuit Notes

Timing resistor R may be adjusted to any value between 10 K and 50 M to obtain a frequency range from 400 kHz to 100 Hz. Returning the timing resistor to the collector of Q1 ensures that Q1 draws its base current only from the timing capacitor Ct. The timing capacitor recharges when the transistors are off, to a voltage equal to the base emitter voltage of Q2 plus the base emitter drops of Q1 and Q2. The transistors then start into conduction. Capacitor Cs is used to speed up the transition. A suitable value would be in the region of 100 pF.

WIEN BRIDGE OSCILLATOR

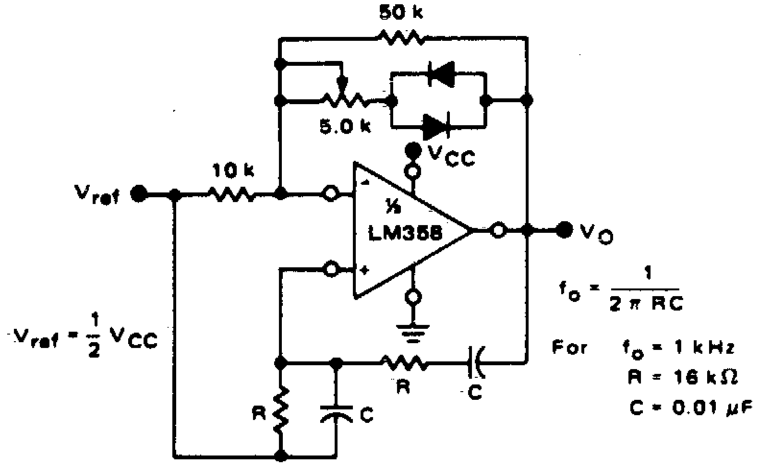


Fig. 7-16

WIEN BRIDGE SINE WAVE OSCILLATOR

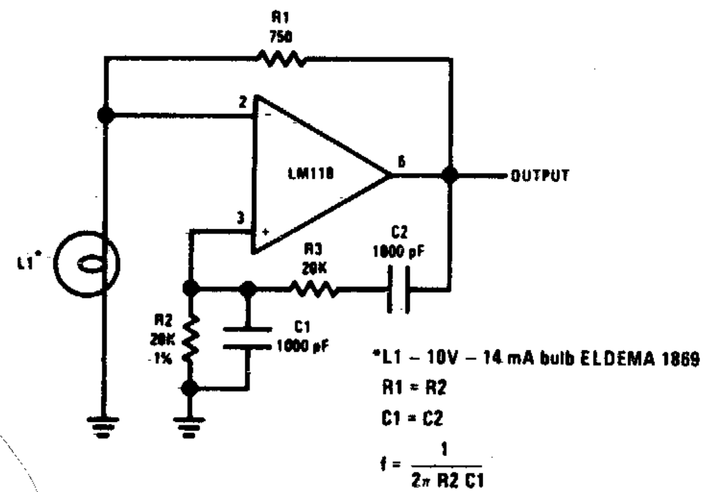


Fig. 7-17

8

Audio Power Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW COST 20 W AUDIO AMPLIFIER

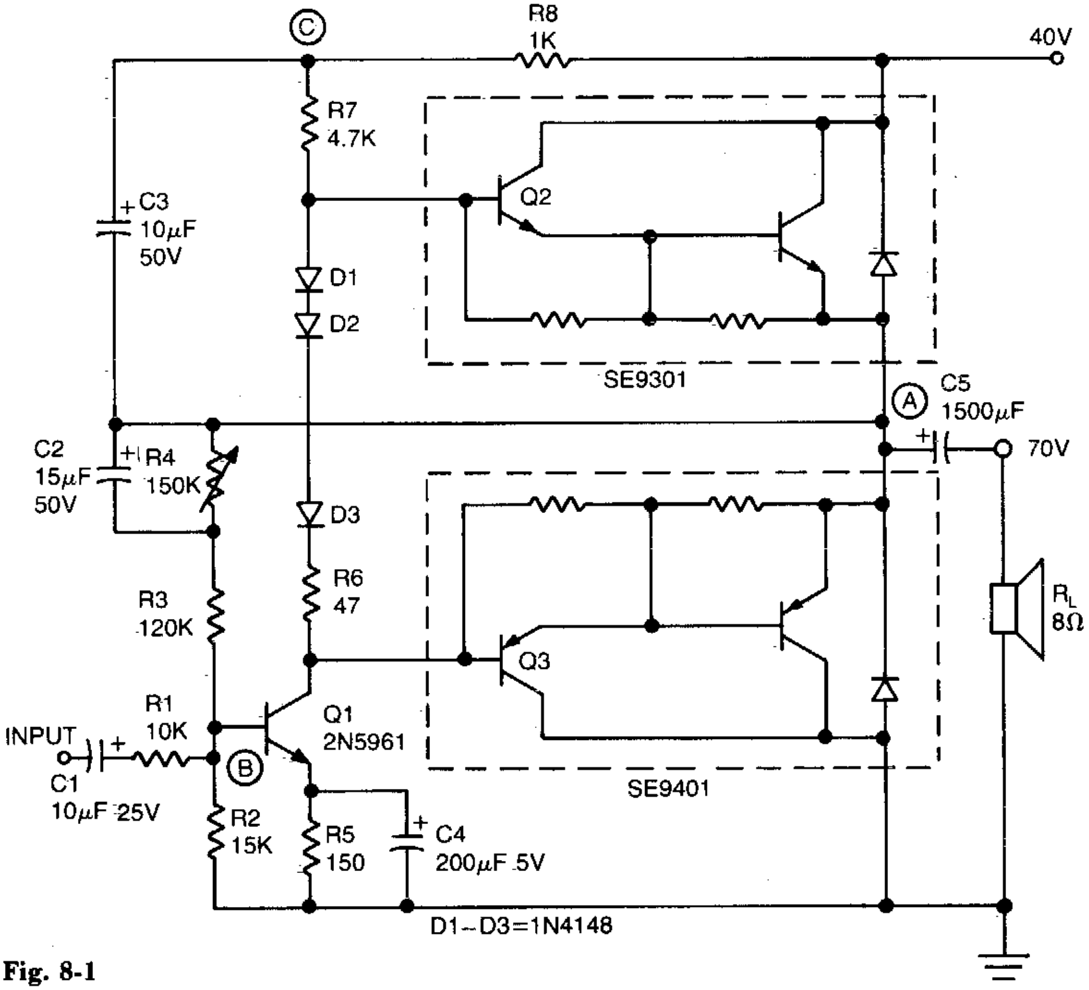


Fig. 8-1

Circuit Notes

This simple inexpensive audio amplifier can be constructed using a couple of TO-220 monolithic Darlington transistors for the push-pull output stage. Frequency response is flat within 1 dB from 30 Hz to 200 kHz with typical harmonic distortion below 0.2%. The amplifier requires only 1.2 V_{rms} for a full 20 W output into an 8 ohm load. Only one other transistor is needed, the TO-92 low-noise high-gain 2N5961 (Q1), to provide voltage gain for driving the output Darlingtons. Its base (point B) is the tie point for ac and dc feedback as well as for the signal input. Input resistance is 10 K. The center voltage at point A is set by adjusting resistor R4. A bootstrap circuit boosts the collector supply voltage of Q1 (point C) to ensure sufficient drive voltage for Q2. This also provides constant voltage across R7, which therefore acts as a current source and, together with diodes D1-D3, reduces low-signal crossover distortion.

75 WATT AUDIO AMPLIFIER WITH LOAD LINE PROTECTION

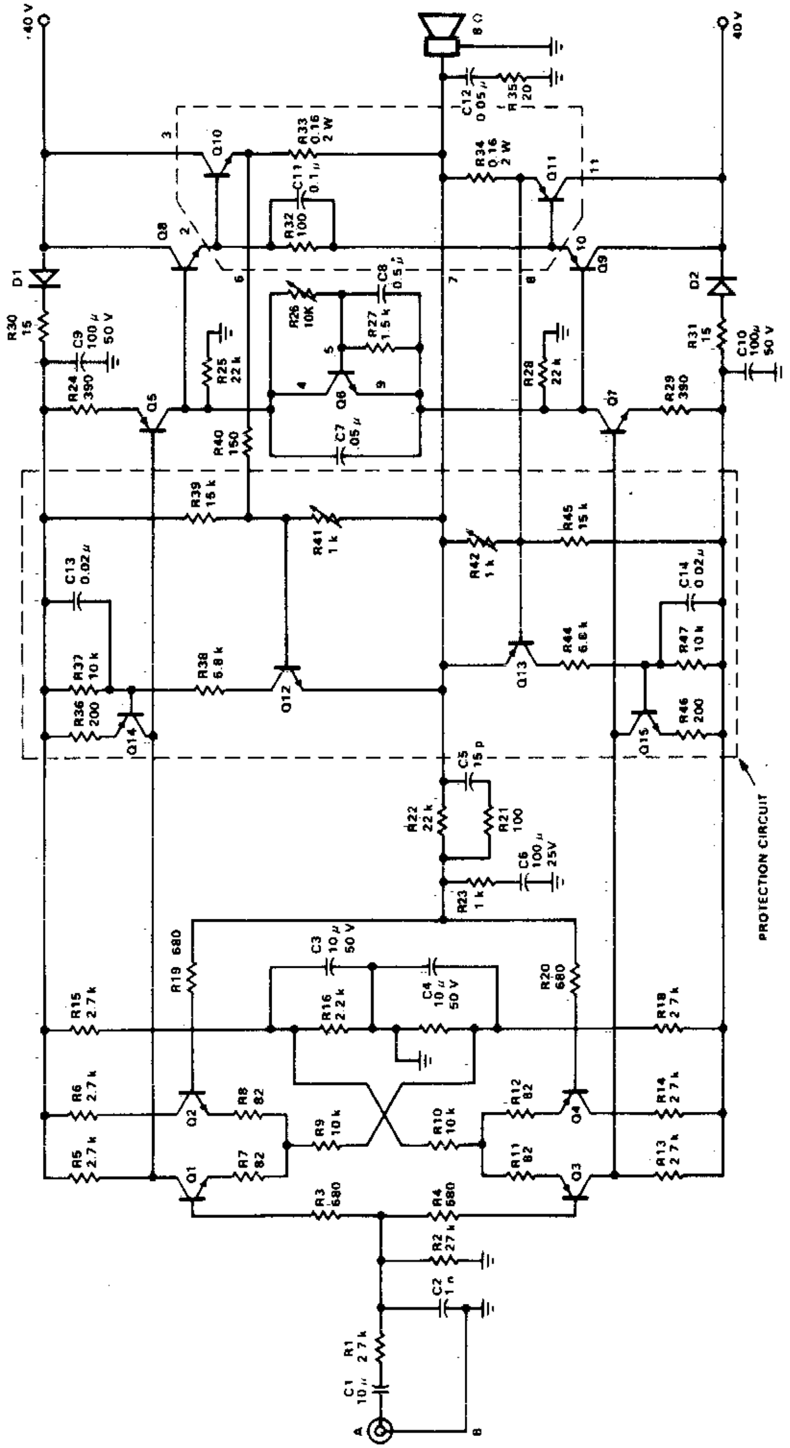


Fig. 8-2

BRIDGE AMPLIFIER

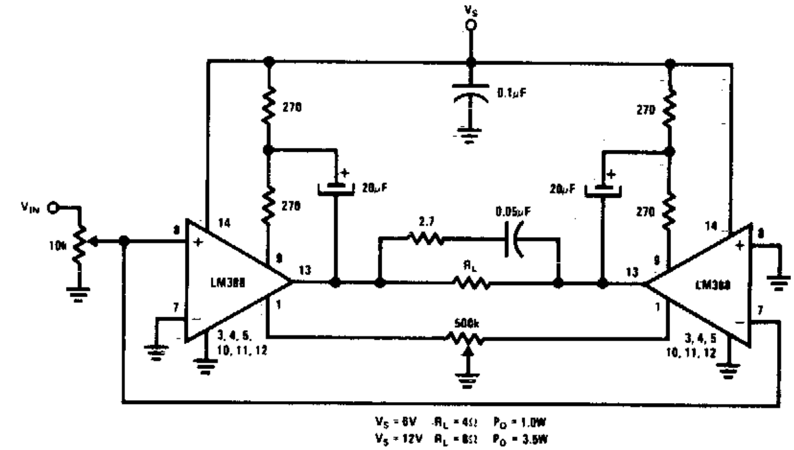


Fig. 8-3

Circuit Notes

This circuit is for low voltage applications requiring high power outputs. Output power levels of 1.0 W into 4 ohm from 6 V and 3.5 V into 8 ohm from 12 V are typical. Coupling capacitors are not necessary since the output dc levels will be within a few tenths of a volt of each other. Where critical matching is required the 500 K potentiometer is added and adjusted for zero dc current flow through the load.

NONINVERTING AMPLIFIER USING SINGLE SUPPLY

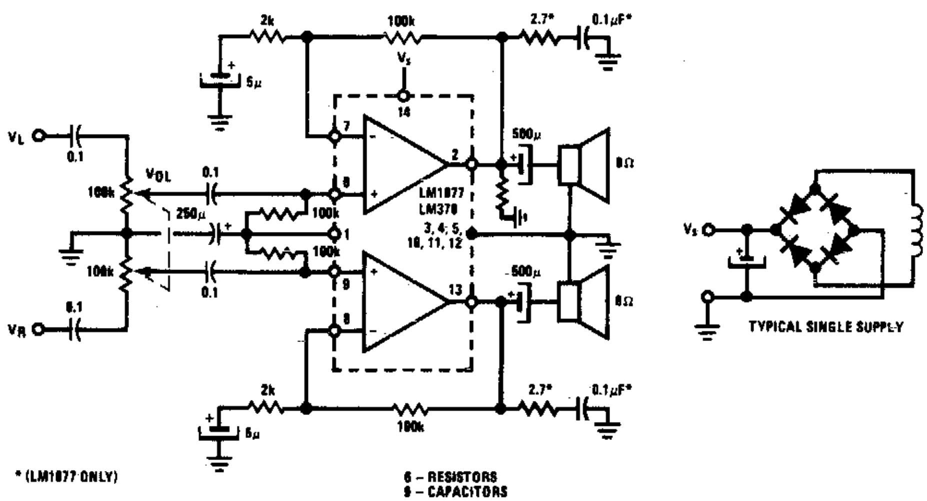


Fig. 8-4

NONINVERTING AMPLIFIER USING SPLIT SUPPLY

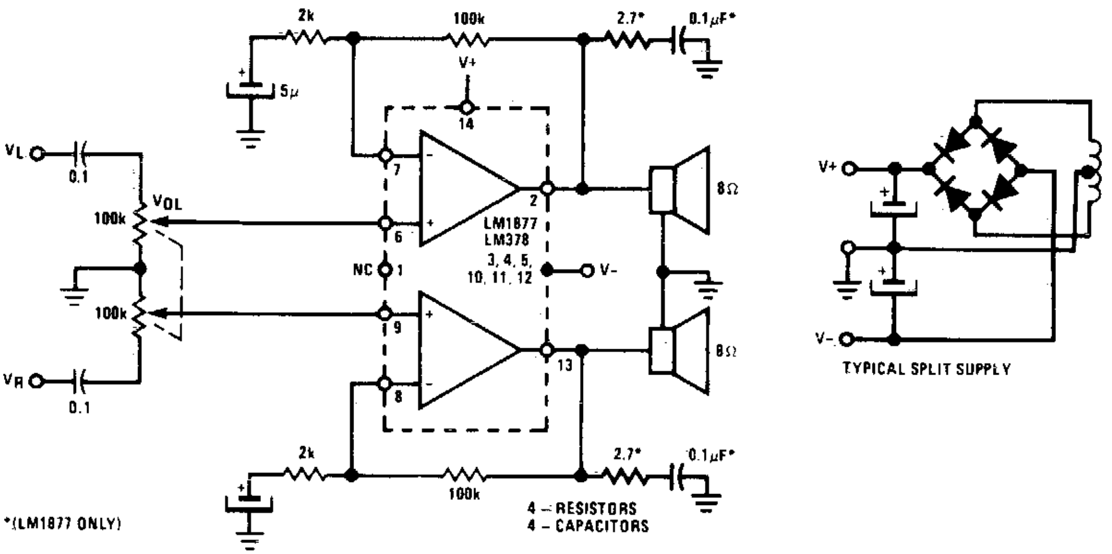
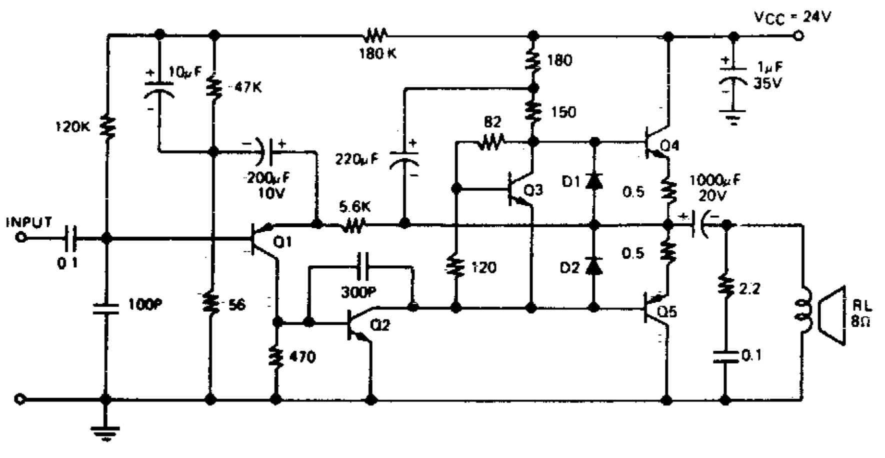


Fig. 8-5

6 W,8 'OHM OUTPUT TRANSFORMERLESS AMPLIFIER



Q1	NB021EY
Q2	NB211YY
Q3	NR001E
Q4	NA41U
Q5	NA42U

Fig. 8-6

12 W LOW-DISTORTION POWER AMPLIFIER

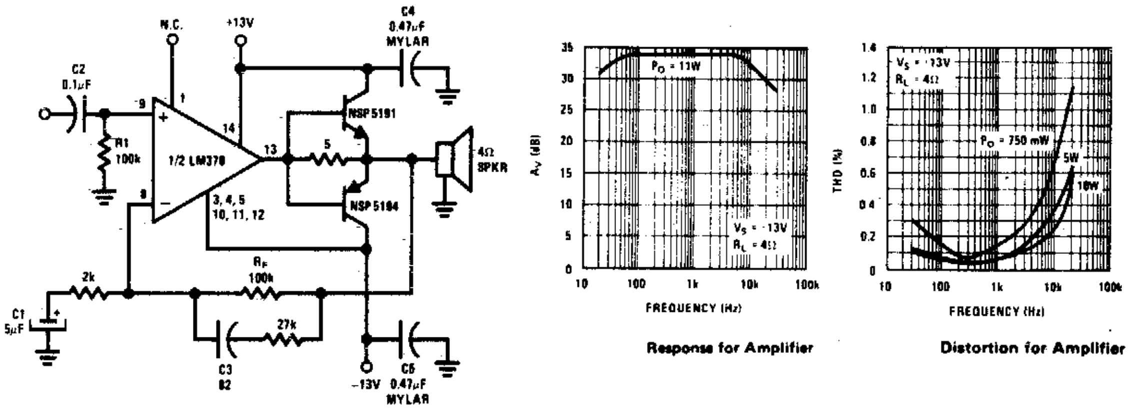


Fig. 8-7

10 W POWER AMPLIFIER

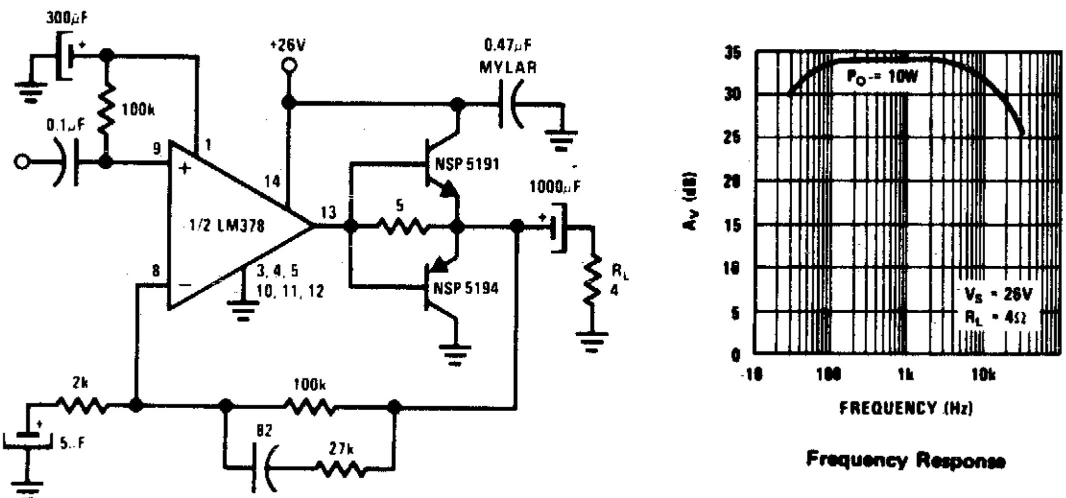


Fig. 8-8

STEREO AMPLIFIER WITH $A_V = 200$

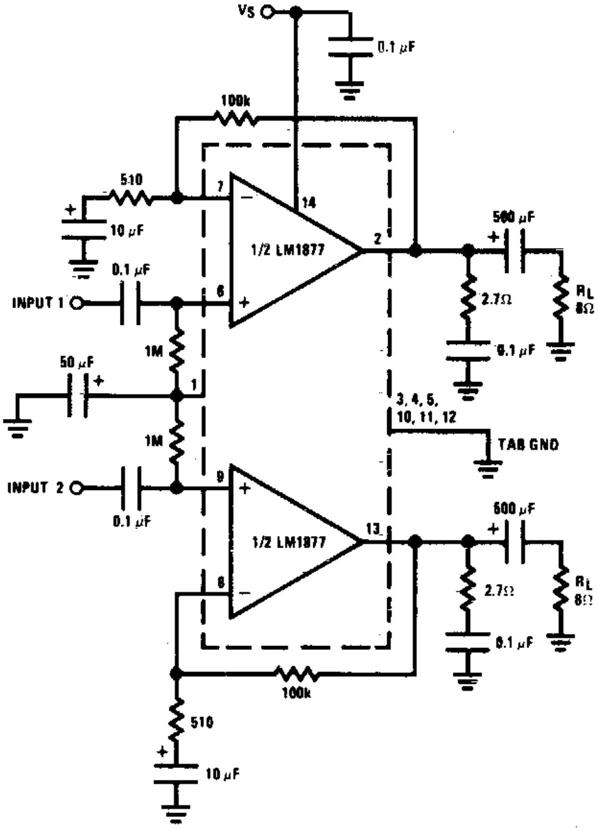
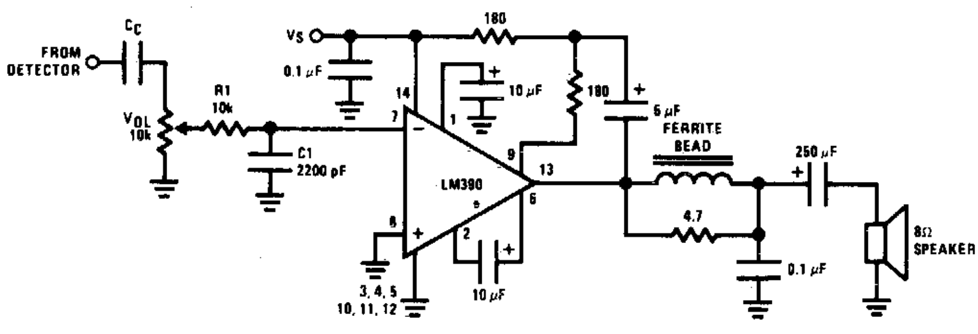


Fig. 8-9

AM RADIO POWER AMPLIFIER



Note 1: Twist supply lead and supply ground very tightly.

Note 2: Twist speaker lead and ground very tightly.

Note 3: Ferrite bead is Ferroxcube K5-001-001/3B with 3 turns of wire.

Note 4: R1C1 band limits input signals.

Note 5: All components must be spaced very close to IC.

Fig. 8-10

470 mW COMPLEMENTARY-SYMMETRY AUDIO AMPLIFIER

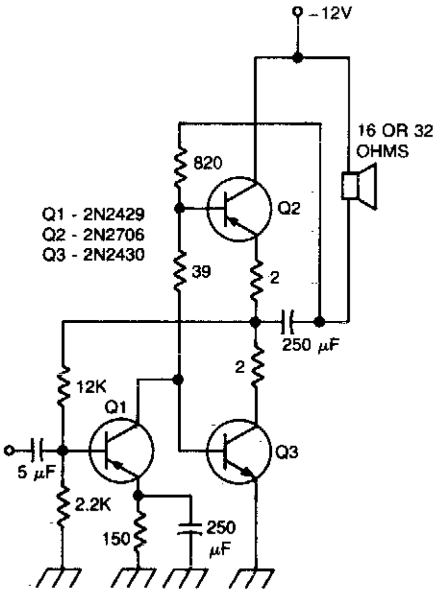


Fig. 8-11

Circuit Notes

This circuit has less than 2% distortion and is flat within 3 dB from 15 Hz to 130 kHz.

NOVEL LOUDSPEAKER COUPLING CIRCUIT

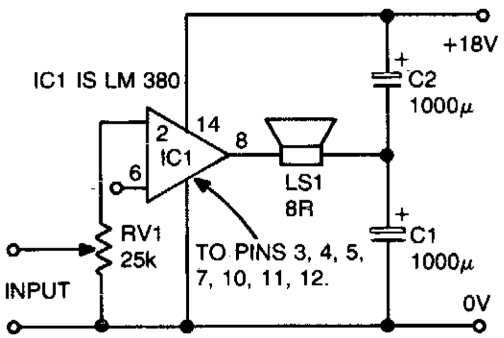


Fig. 8-12

Circuit Notes

The ground side of the speaker is connected to the junction of two equal high value capacitors (1000 μF is typical) across the supply. The amplifier output voltage will be $V_s/2$, and so will the voltage across C1 (if C1 and C2 are equal); so as the supply voltage builds up, the dc voltage across the speaker will remain zero, eliminating the switch-on surge. C1 and C2 will also provide supply smoothing. The circuit is shown with the LM380, but could be applied to any amplifier circuit, providing that the dc voltage at the output is half the supply voltage.

NONINVERTING AC POWER AMPLIFIER

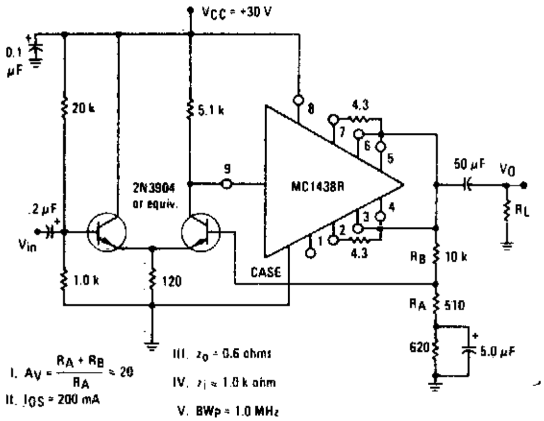


Fig. 8-13

NONINVERTING POWER AMPLIFIER

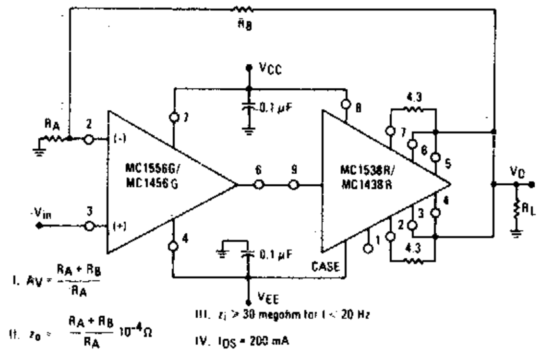


Fig. 8-15

INVERTING POWER AMPLIFIER

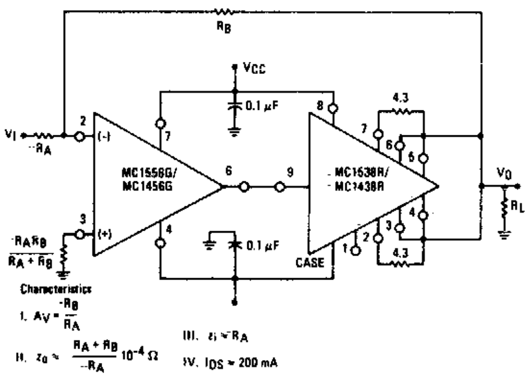


Fig. 8-14

4 W BRIDGE AMPLIFIER

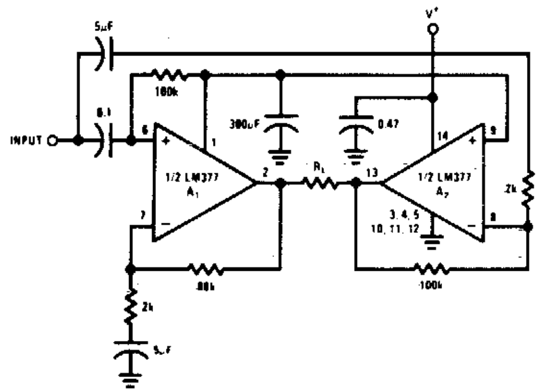
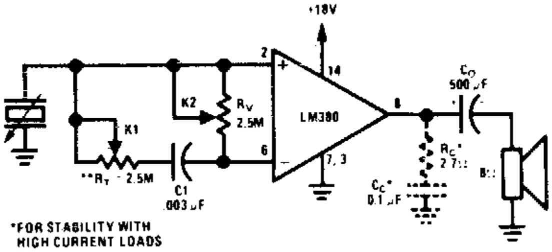


Fig. 8-16

PHONO AMPLIFIER WITH "COMMON MODE" VOLUME AND TONE CONTROL



*FOR STABILITY WITH HIGH CURRENT LOADS

**AUDIO TAPE POTENTIOMETER (10% OF R_T AT 50% ROTATION)

Fig. 8-17

PHONOGRAPH AMPLIFIER (CERAMIC CARTRIDGE)

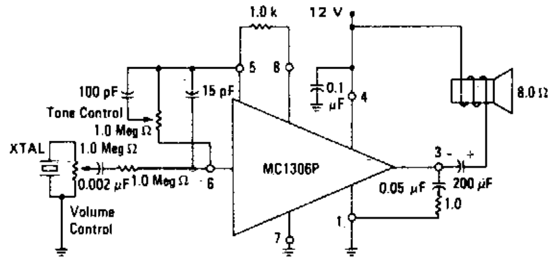


Fig. 8-19

PHONO AMPLIFIER

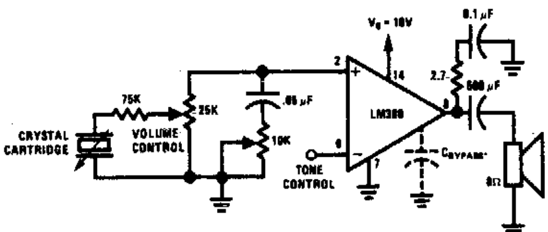


Fig. 8-18

INVERTING UNITY GAIN AMPLIFIER

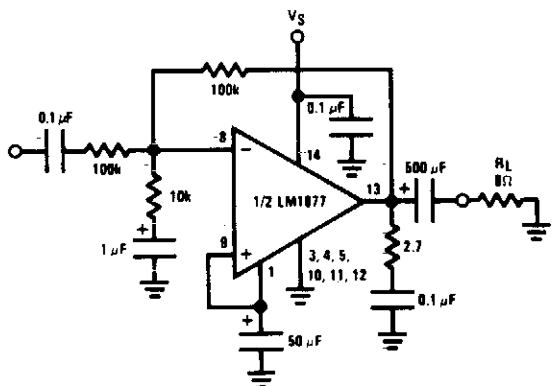
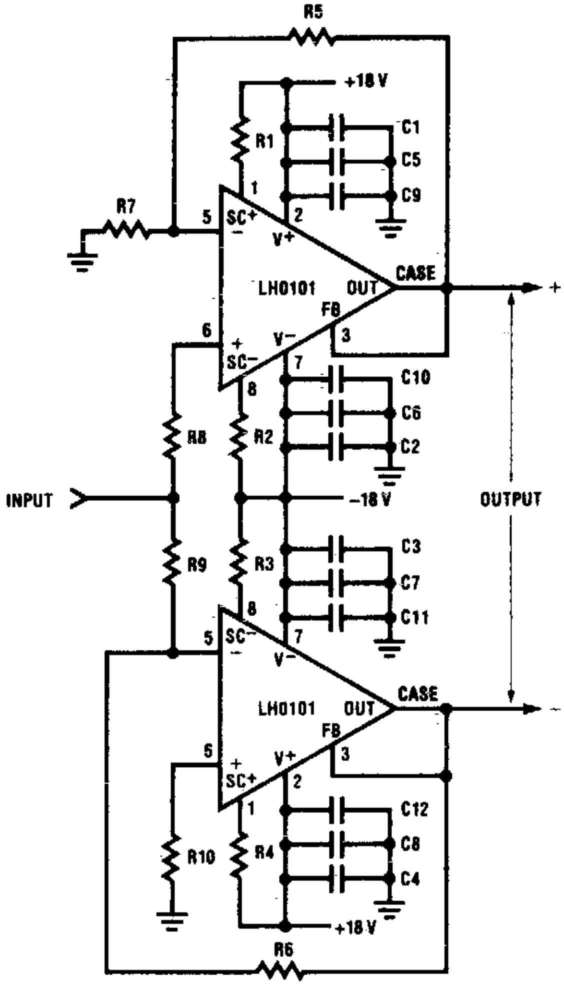
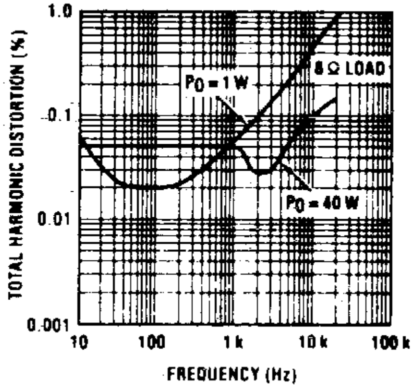


Fig. 8-20

BRIDGE AUDIO POWER AMPLIFIER



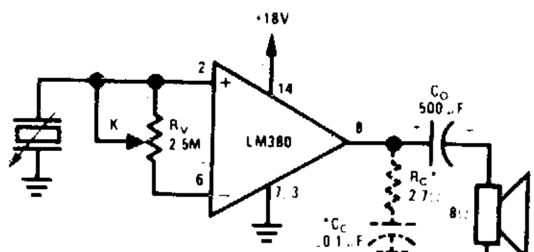


Total Harmonic Distortion vs. Frequency of Bridge Power Amplifier

R1-R4	CURRENT LIMIT RESISTOR	0.15 Ω 2 W
R5	FEEDBACK RESISTOR	5 kΩ
R6	FEEDBACK RESISTOR	15 kΩ
R7-R10	INPUT RESISTORS	10 kΩ
C1-C4	BYPASS CAPACITORS	47 μF 25 V ELECTROLYTIC
C5-C8	BYPASS CAPACITORS	10 μF 25 V TANTALUM
C9-C12	BYPASS CAPACITORS	0.1 μF 25V CERAMIC

Fig. 8-21

PHONO AMPLIFIER



*FOR STABILITY WITH HIGH CURRENT LOADS

Circuit Notes

Used when maximum input impedance is required or the signal attenuation of the voltage divider volume control is undesirable.

Fig. 8-22

HIGH SLEW RATE POWER OP AMP/AUDIO AMP

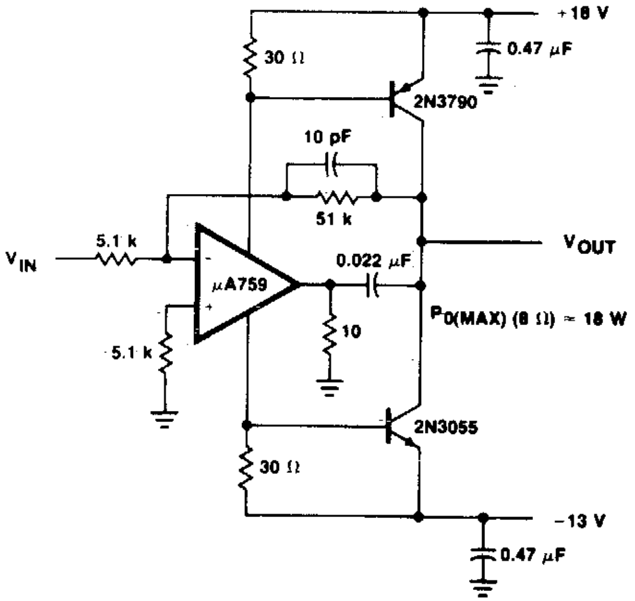


Fig. 8-23

Features

- High Slew Rate 9 V/μs
- High 3 dB Power Bandwidth 85 kHz
- 18 Watts Output Power Into an 8 Ω Load.
- Low Distortion — .2%, 10 VRMS, 1 kHz Into 8 Ω

16 W BRIDGE AMPLIFIER

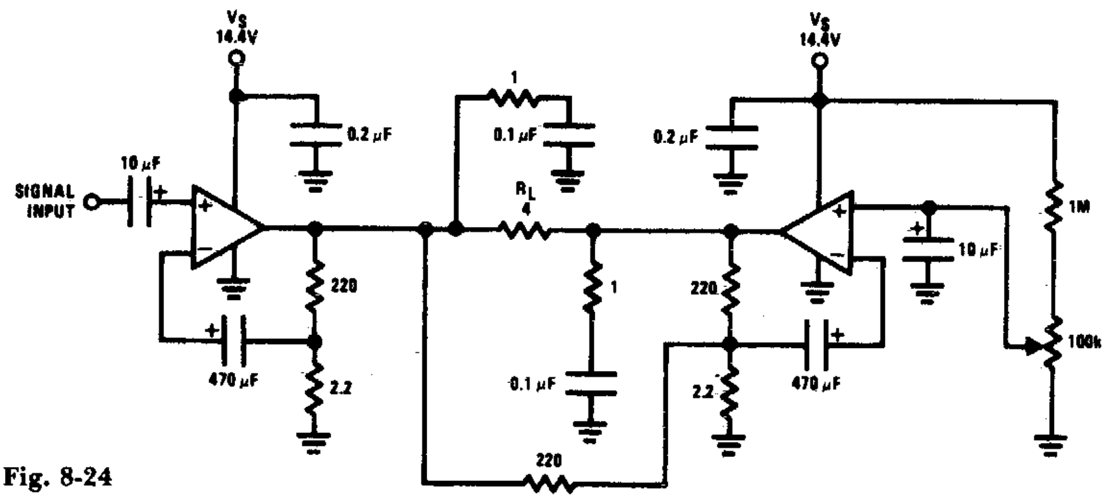


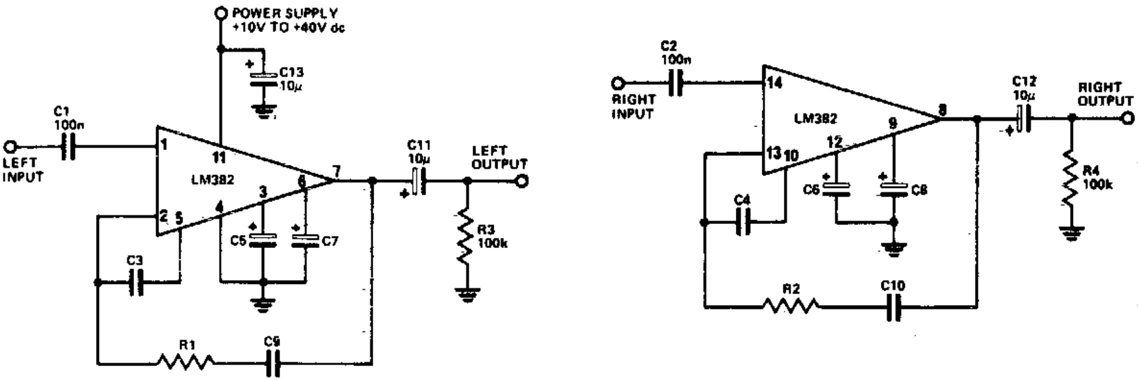
Fig. 8-24

9

Audio Signal Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

GENERAL PURPOSE PREAMPLIFIER



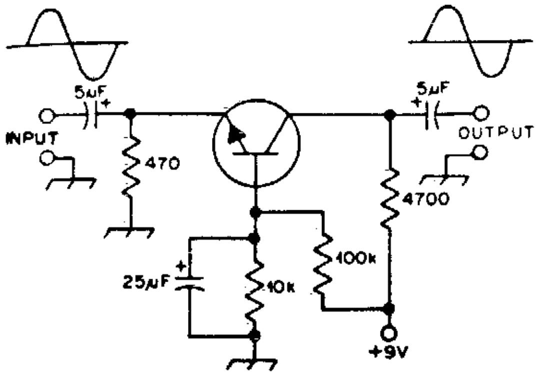
FUNCTION	C3, 4	C5, 6	C7, 8	C9, 10	R1, 2
Phono preamp (RIAA)	330n	10μF	10μF	1n5	1k
Tape preamp (NAB)	68n	10μF	10μF	–	–
Flat 40dB gain	–	–	10μF	–	–
Flat 55dB gain	–	10μF	–	–	–
Flat 80dB gain	–	10μF	10μF	–	–

Fig. 9-1

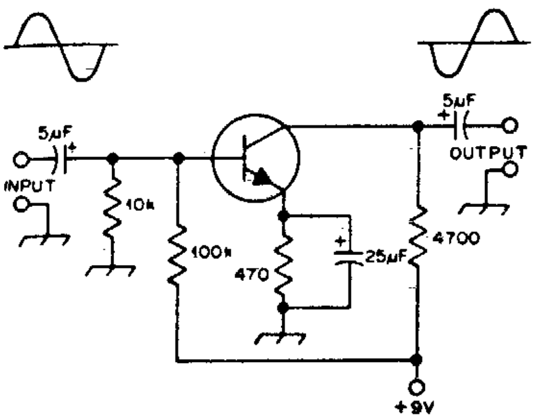
Circuit Notes

Not much can be said about how the LM382 works as most of the circuitry is contained within the IC. Most of the frequency-determining components are on the chip—only the capacitors are mounted externally. The LM382 has the convenient characteristic of rejecting ripple on the supply line by about 100 dB, thus greatly reducing the quality requirment for the power supply.

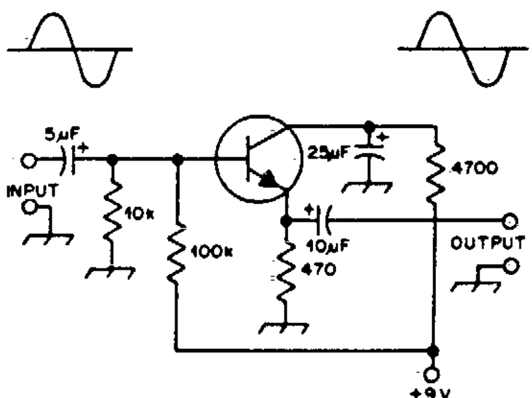
BASIC TRANSISTOR AMPLIFIER CIRCUITS



COMMON BASE



COMMON EMITTER



COMMON COLLECTOR

Circuit Notes

Typical component values are given for use at audio frequencies, where these circuits are used most often. The input and output phase relationships are shown.

Fig. 9-2

ELECTRONIC BALANCED INPUT MICROPHONE AMPLIFIER

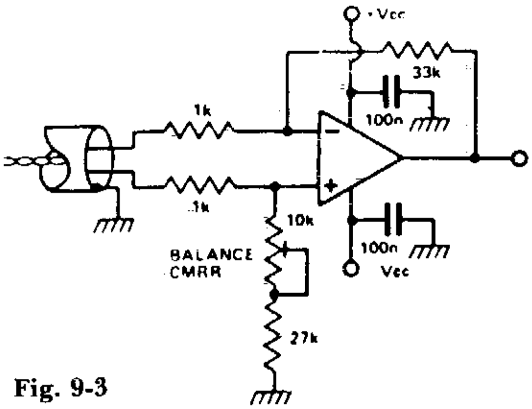


Fig. 9-3

Circuit Notes

It is possible to simulate the balanced performance of a transformer electronically with a different amplifier. By adjusting the presets, the resistor ratio can be balanced so that the best CMRR is obtained. It is possible to get a better CMRR than from a transformer. Use a RC4136 which is a quad low noise op amp.

TRANSDUCER AMPLIFIER

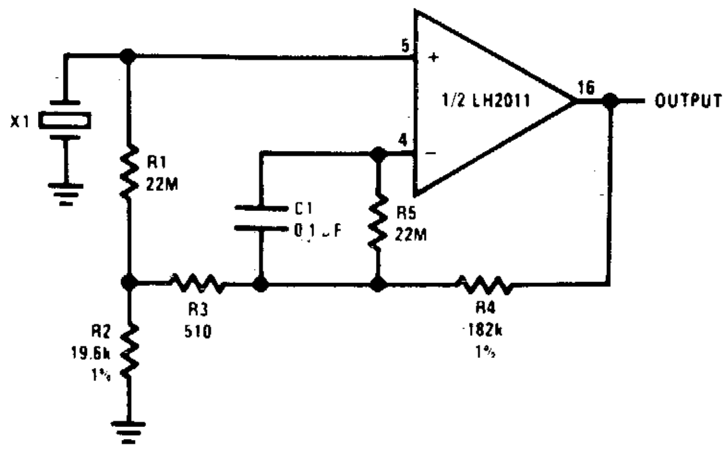


Fig. 9-4

$$R_{IN} = R1\left(1 + \frac{R2}{R3}\right) \quad A_V = \frac{R2 + R3 + R4}{R2 + R3}$$

Circuit Notes

This circuit is high-input-impedance ac amplifier for a piezoelectric transducer. Input resistance is 880 M, and a gain of 10 is obtained.

ULTRA-HIGH GAIN AUDIO AMPLIFIER

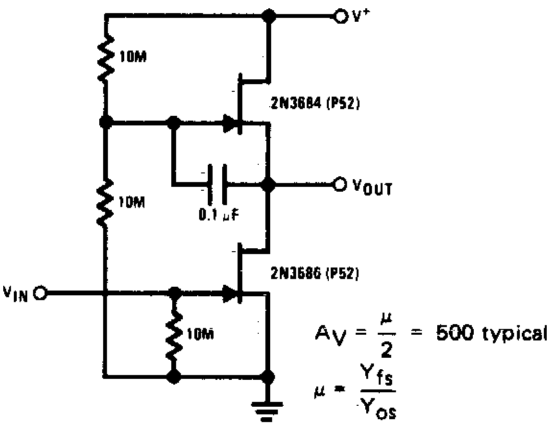


Fig. 9-5

Circuit Notes

Sometimes called the JFET μ-amp, this circuit provides a very low power, high gain amplifying function. Since μ of a JFET increases as drain current decreases, the lower drain current is, the more gain you get. Input dynamic range is sacrificed with increasing gain, however.

MICROPHONE AMPLIFIER

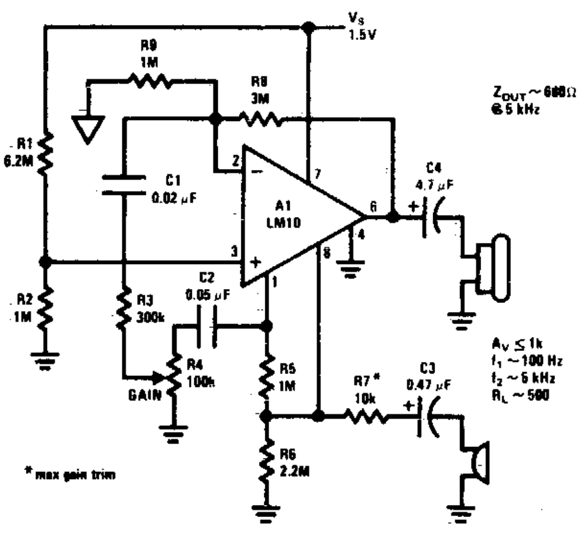


Fig. 9-6

Circuit Notes

This circuit operates from a 1.5 Vdc source.

TRANSFORMERLESS (BALANCE INPUTS) MICROPHONE PREAMP

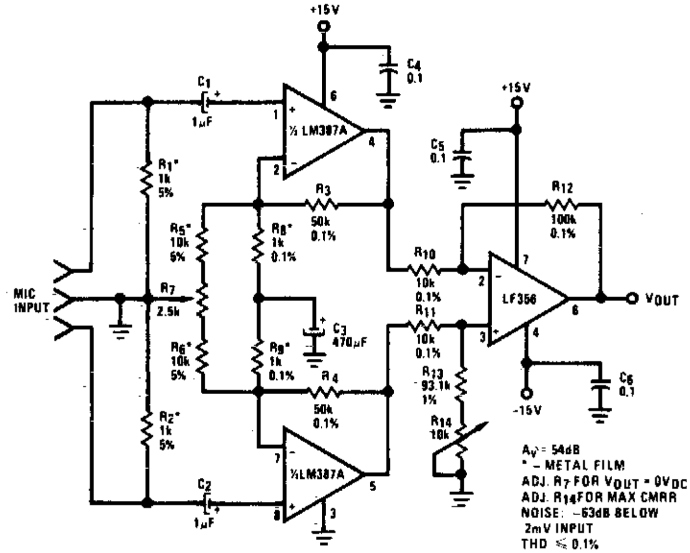
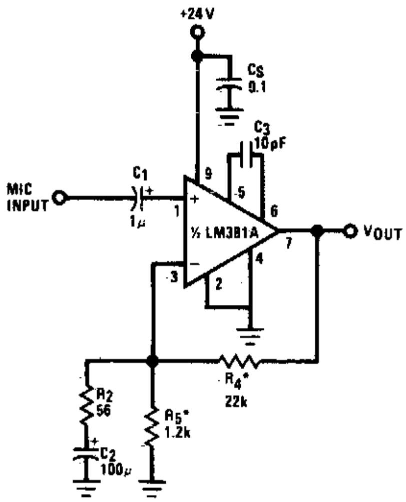


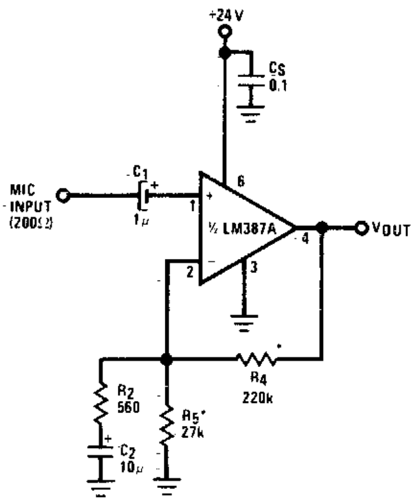
Fig. 9-7

TRANSFORMERLESS MICROPHONE PREAMPS (UNBALANCED INPUTS)



A_V = 52dB
* – METAL FILM
NOISE: –69dB BELOW
2mV
THD 0.1%

(a) LM381A S. E. Bias



A_V = 52dB
* – METAL FILM
NOISE: –67dB BELOW
2mV
THD 0.1%

(b) LM387A

Fig. 9-8

MAGNETIC PICKUP PHONO PREAMPLIFIER

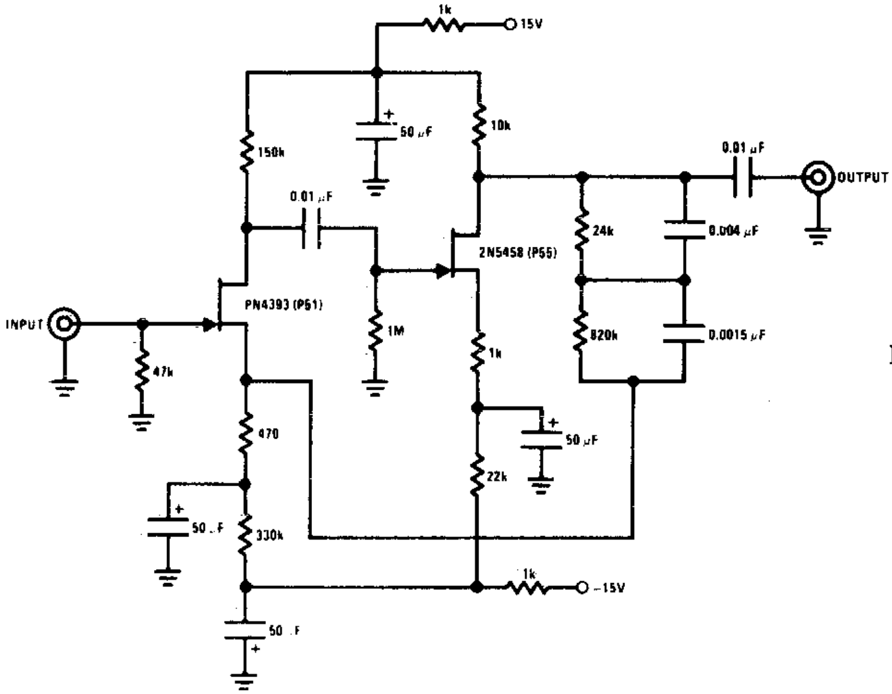


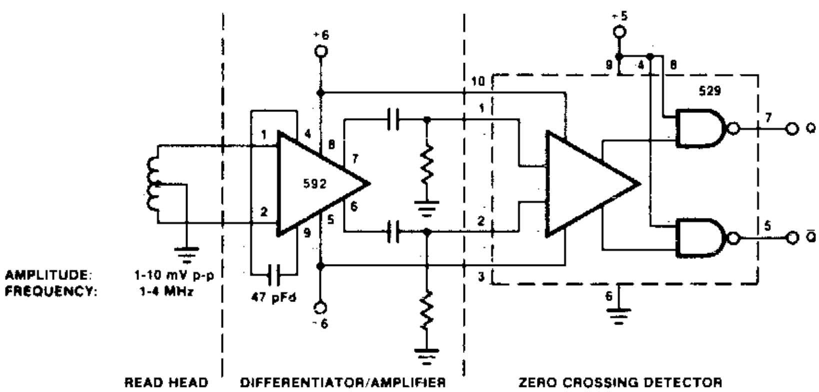
Fig. 9-9

Circuit Notes

This preamplifier provides proper loading to a reluctance phono cartridge. It provides approximately 35 dB of gain at 1 kHz (2.2 mV input for 100 mV output). It features (S + N)/N ratio of better than −70 dB (referenced to 10 mV input at 1 kHz) and has a dynamic range of 84 dB (referenced to 1 kHz). The feedback provides for RIAA equalization.

DISC/TAPE PHASE MODULATED READBACK SYSTEMS

Fig. 9-10



TWO-POLE FAST TURN-ON NAB TAPE PREAMPLIFIER

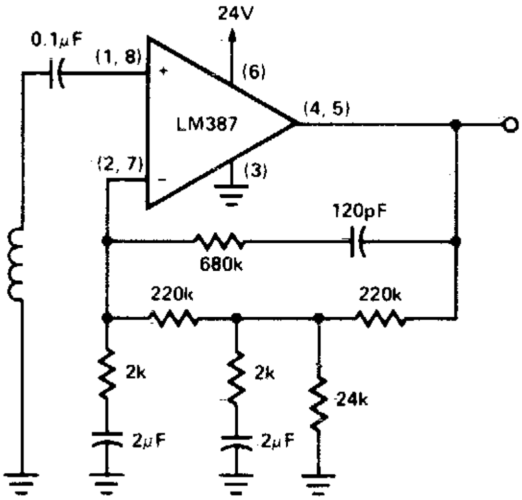


Fig. 9-11

LM382 PHONO PREAMPLIFIER (RIAA)

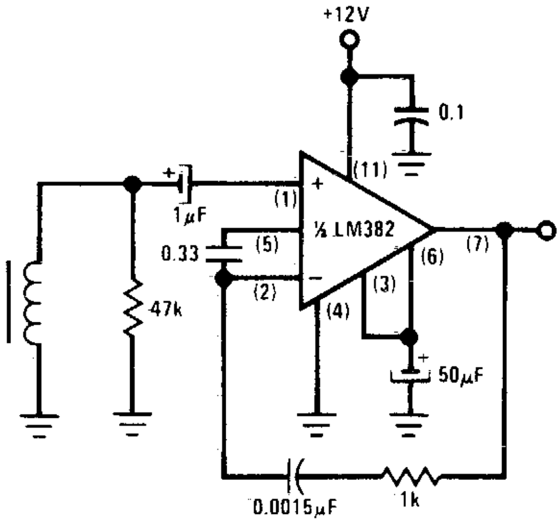


Fig. 9-13

TAPE PREAMPLIFIER (NAB EQUALIZATION)

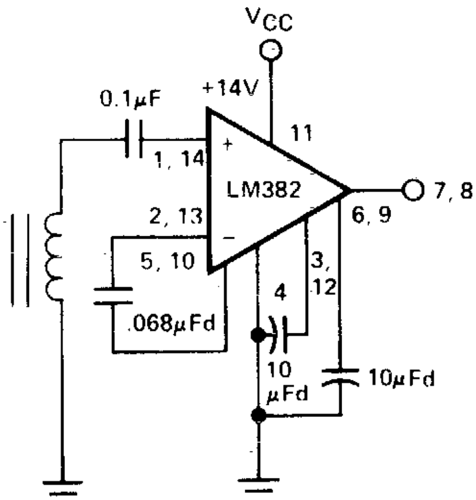


Fig. 9-12

TAPE RECORDING AMPLIFIER

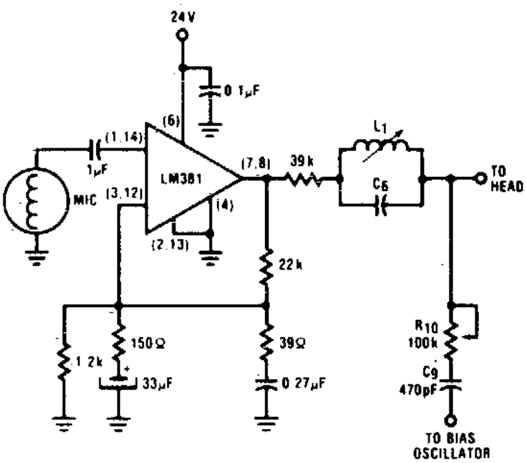


Fig. 9-14

MAGNETIC PHONO PREAMPLIFIER

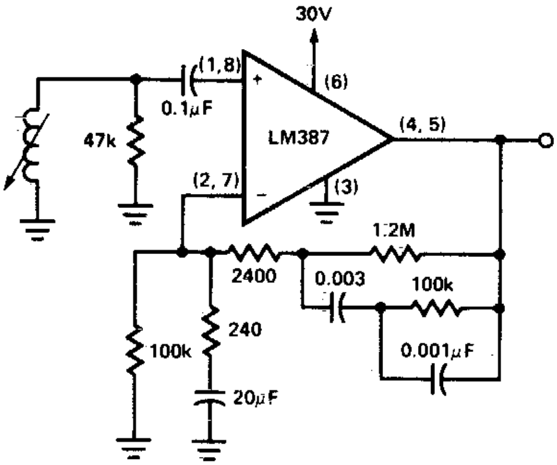


Fig. 9-15

REMOTE AMPLIFIER

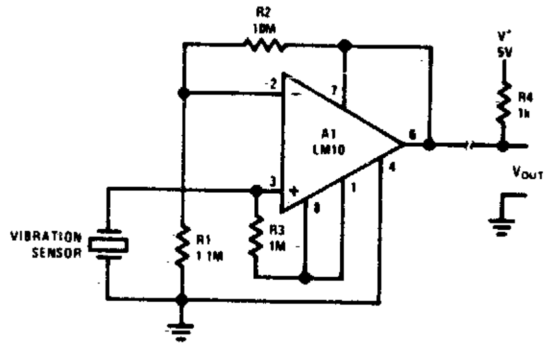
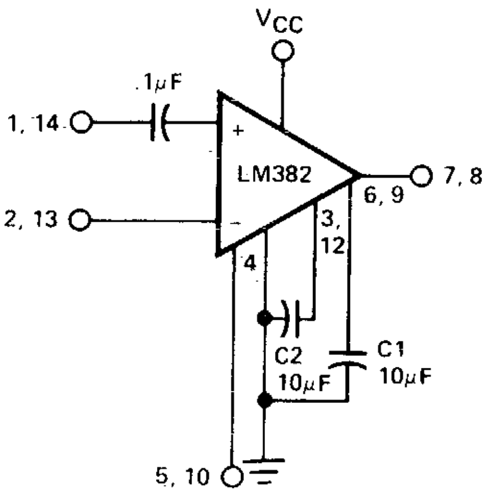


Fig. 9-17

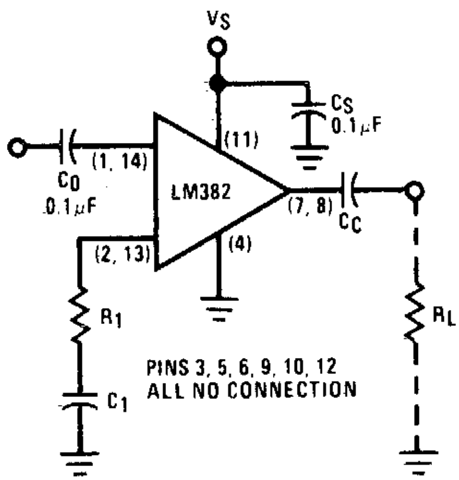
PHONO PREAMP (RIAA EQUALIZATION)



CAPACITOR	GAIN
C1 Only	40dB
C2 Only	55dB
C1 & C2	80dB

Fig. 9-16

ADJUSTABLE GAIN NONINVERTING AMPLIFIER



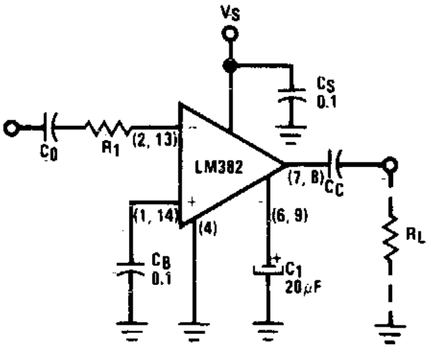
$$\text{GAIN} = 1 + \frac{267k}{R_1}$$

$$C_1 = \frac{1}{2\pi f_0 R_1}$$

f_0 = LOW FREQUENCY -3dB CORNER

Fig. 9-18

HIGH GAIN INVERTING AC AMPLIFIER



$$\text{GAIN} = -\frac{5.1 \times 10^6}{R_1}$$

$$C_0 = \frac{1}{2\pi f_0 R_1}$$

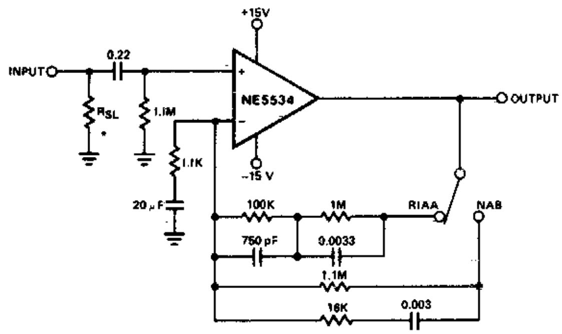
f_0 = LOW FREQUENCY -3dB CORNER $(C_C R_L >> C_0 R_1)$

INPUT IMPEDANCE = R_1

PINS 3, 5, 10, 12 NOT USED

Fig. 9-19

PREAMPLIFIER WITH RIAA/NAB COMPENSATION



*Select to provide specified transducer loading.
Output Noise ≅ 0.8mV rms (with input shorted)

All resistor values are in ohms.

Fig. 9-21

FLAT RESPONSE AMPLIFIER (FIXED GAIN CONFIGURATION)

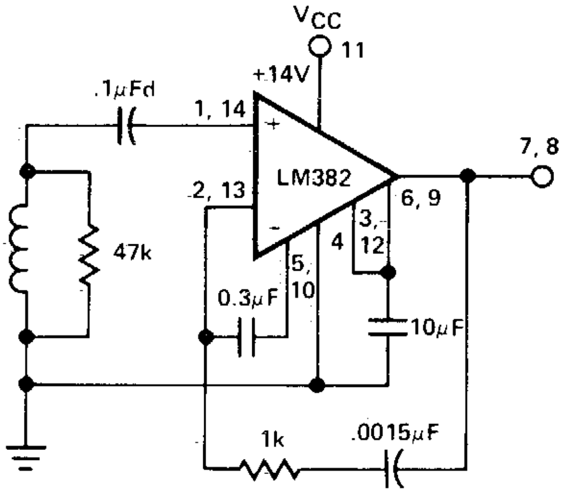


Fig. 9-20

TAPE PLAYBACK AMPLIFIER

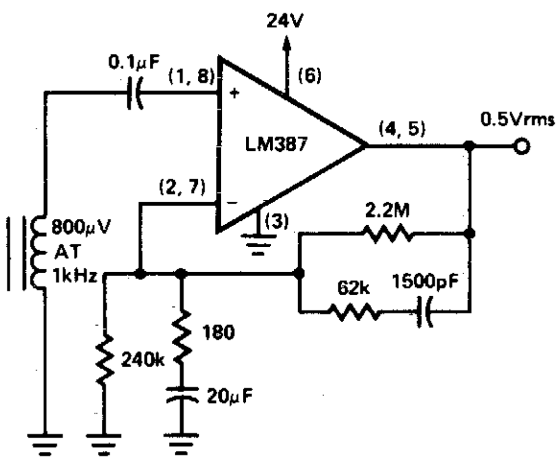


Fig. 9-22

10

Automotive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Gasoline Engine Tachometer
Speed Alarm
Speed Warning Device
Universal Wiper Delay
Courtesy Light Extender
Bargraph Car Voltmeter
Tachometer
High Speed Warning Device
Breaker Point Dwell Meter
Tachometer
Capacitor Discharge Ignition System
Windshield Wiper Control
Auto Battery Current Analyzer
Speed Switch
Windshield Wiper Controller
Windshield Wiper Hesitation Control Unit
Ice Warning and Lights Reminder
Car Battery Monitor
Headlight Delay Unit
Windshield Washer Fluid Watcher
Car Battery Condition Checker
Overspeed Indicator
Sequential Fiasher for Auto Turn Signals
Auto Lights-On Reminder

GASOLINE ENGINE TACHOMETER

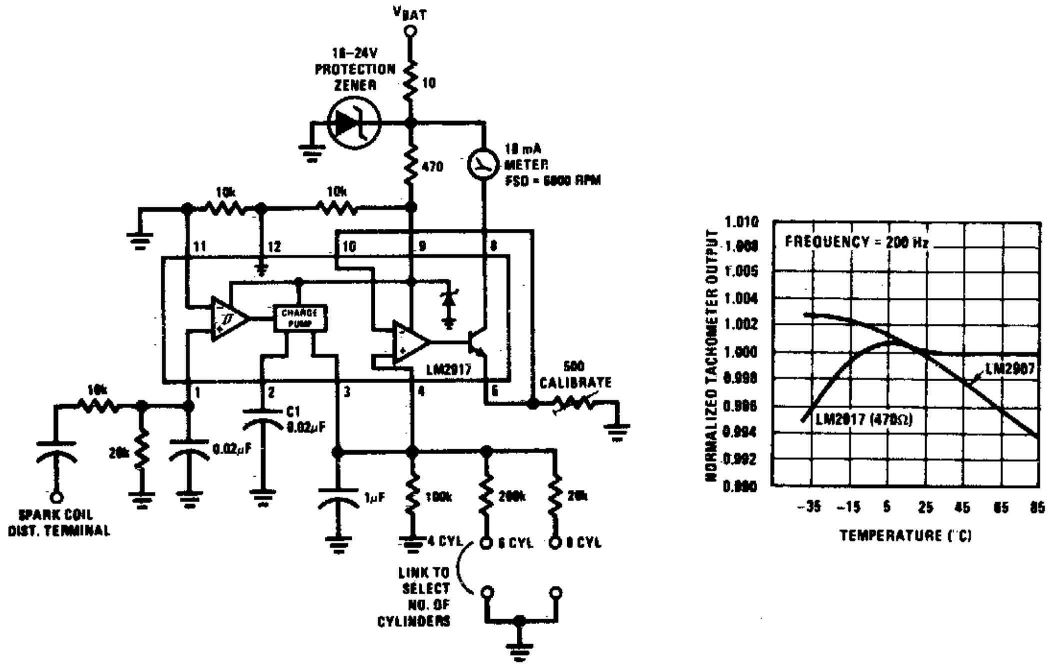


Fig. 10-1

Circuit Notes

This tachometer can be set up for any number of cylinders by linking the appropriate timing resistor as illustrated. A 500 ohm trim resistor can be used to set up final calibration. A protection circuit composed of a 10 ohm resistor and a zener diode is also shown as a safety precaution against the transients which are to be found in automobiles.

SPEED ALARM

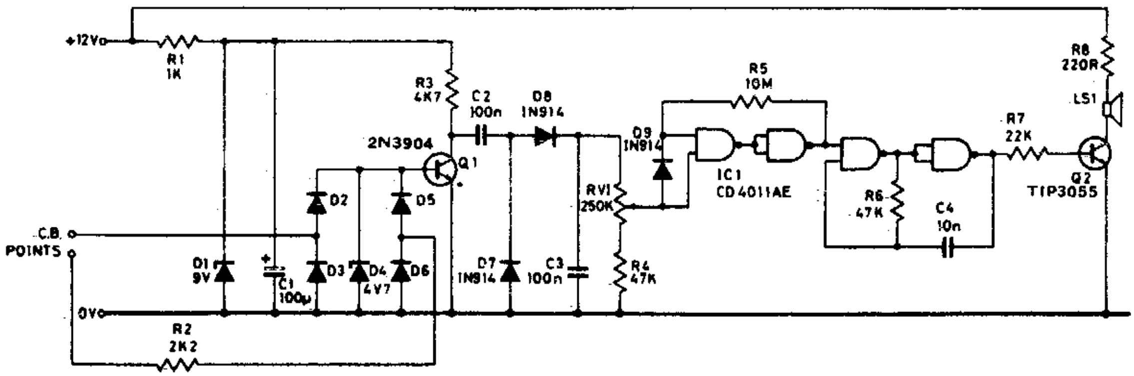
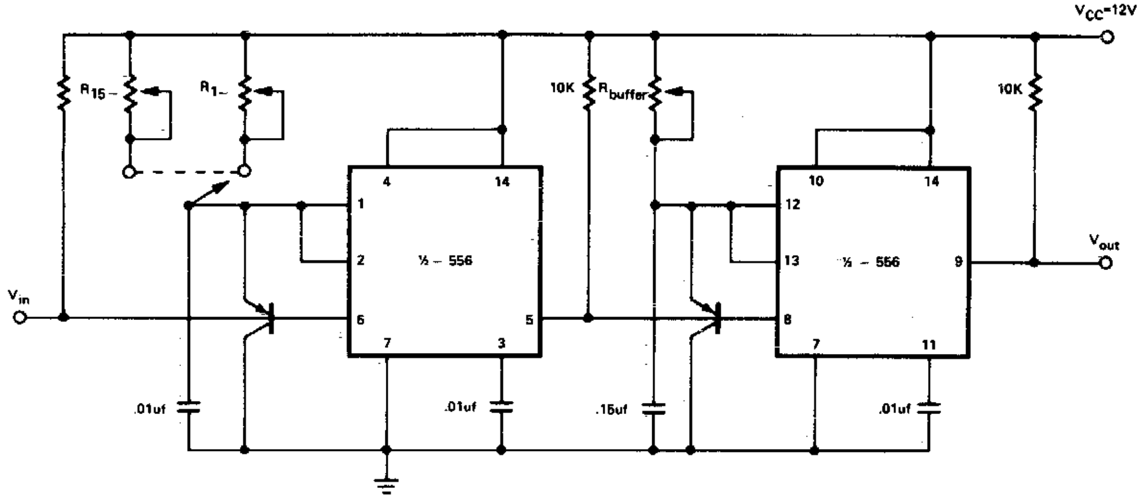


Fig. 10-2

Circuit Notes

Pulses from the distributor points are passed through a current limiting resistor, rectified, and clipped at 4.7 volts. Via Q1 and the diode pump, a dc voltage proportional to engine rpm is presented to RV1; the sharp transfer characteristic of a CMOS gate, assisted by feedback, is used to enable the oscillator formed by the remaining half of the 4011. At the pre-set speed, a nonignorable tone emits from the speaker, and disappears as soon as the speed drops by three or four mph.

SPEED WARNING DEVICE



OPERATING WAVEFORMS

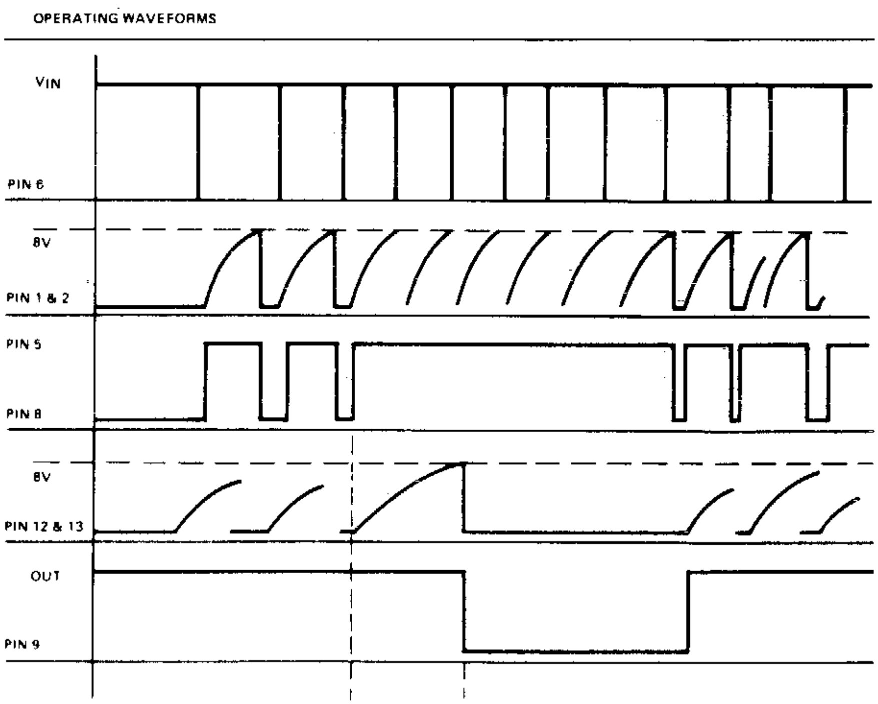


Fig. 10-3

UNIVERSAL WIPER DELAY

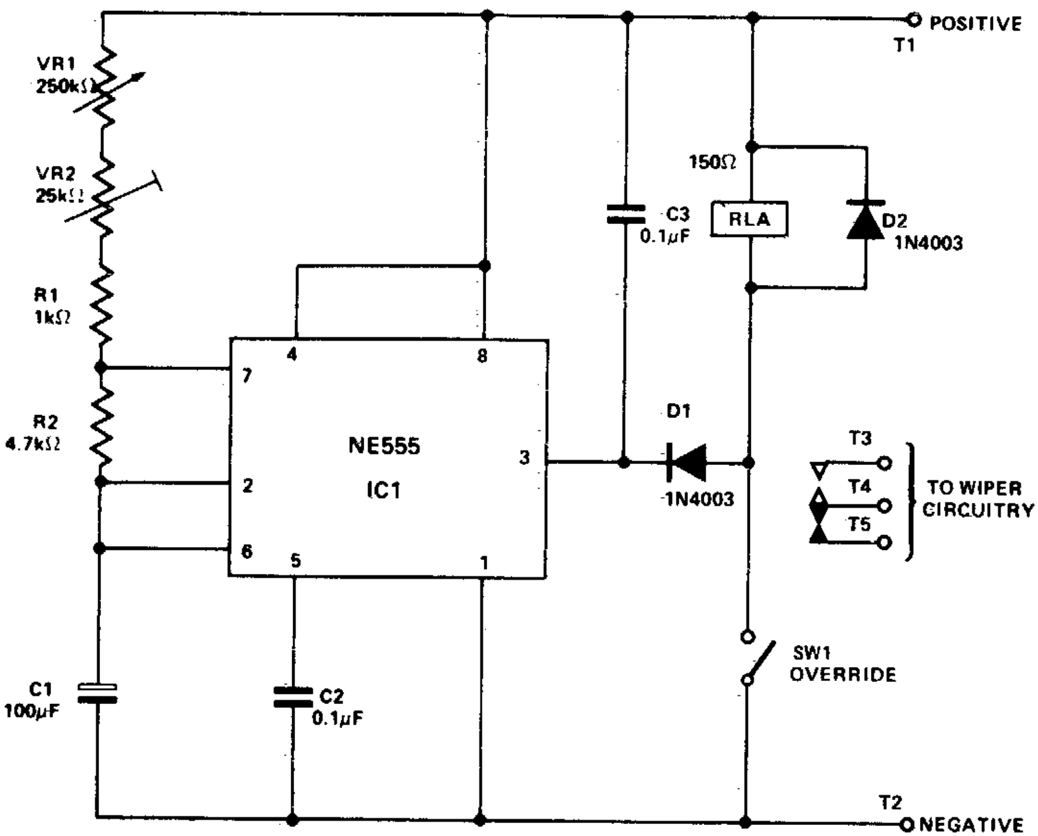


Fig. 10-4

Circuit Notes

IC1 is connected in the astable mode, driving RLA. C3, D1, and D2 prevent spikes from the relay coil and the wiper motor from triggering IC1. VR2 is adjusted to give the minimum delay time required. VR1 is the main delay control and provides a range of from about 1 second to 20 seconds. SW1 is an override switch to hold RLA permanently on (for normal wiper operation). The relay should have a resistance of at least 150 ohms and have heavy duty contacts. The suppression circuit may be needed for the protection of IC1.

COURTESY LIGHT EXTENDER

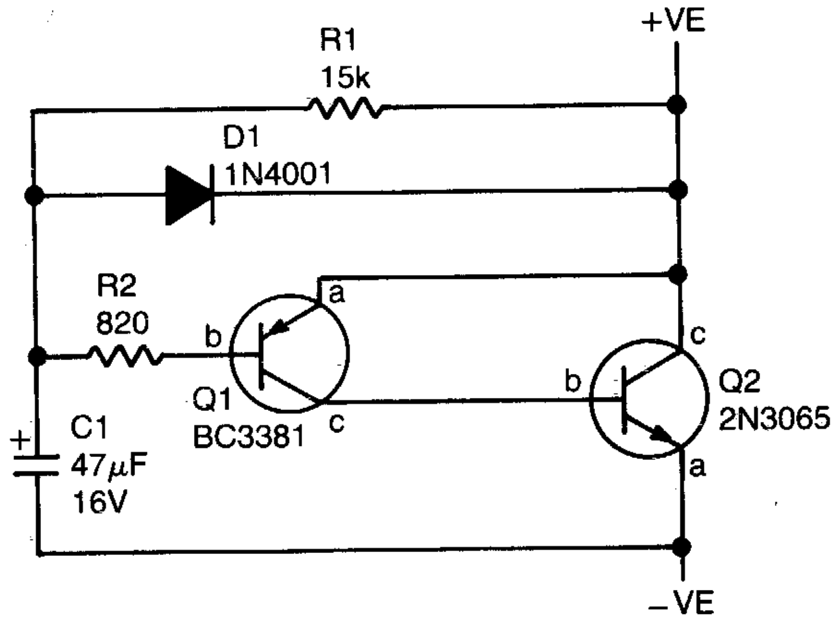


Fig. 10-5

Circuit Notes

Most car door switches are simply single-pole switches, with one side grounded. When the door is opened the switch grounds the other line thus completing the light circuit. In a car where the negative terminal of the battery is connected to the chassis, the negative wire of the unit (emitter of Q2) is connected to chassis the positive wire (case of 2N3055) is connected to the wire going to the switch. In a car having a positive ground system this connection sequence is reversed. When the switch closes (door open), C1 is discharged via D1 to zero volts, and when the switch opens, C1 charges up via R1 and R2. Transistors Q1 and Q2 are connected as an emitter follower (Q2 just buffers Q1) therefore the voltage across Q2 increases slowly as C1 charges. Hence Q2 acts like a low resistance in parallel with the switch and keeps the lights on. The value of C1 is chosen such that a useful light level is obtained for about four seconds; therefore the light decreases until in about 10 seconds it is out completely. With different transistor gains and with variation in current drain due to a particular type of car, the timing may vary but may be simply adjusted by selecting C1.

BARGRAPH CAR VOLTMETER

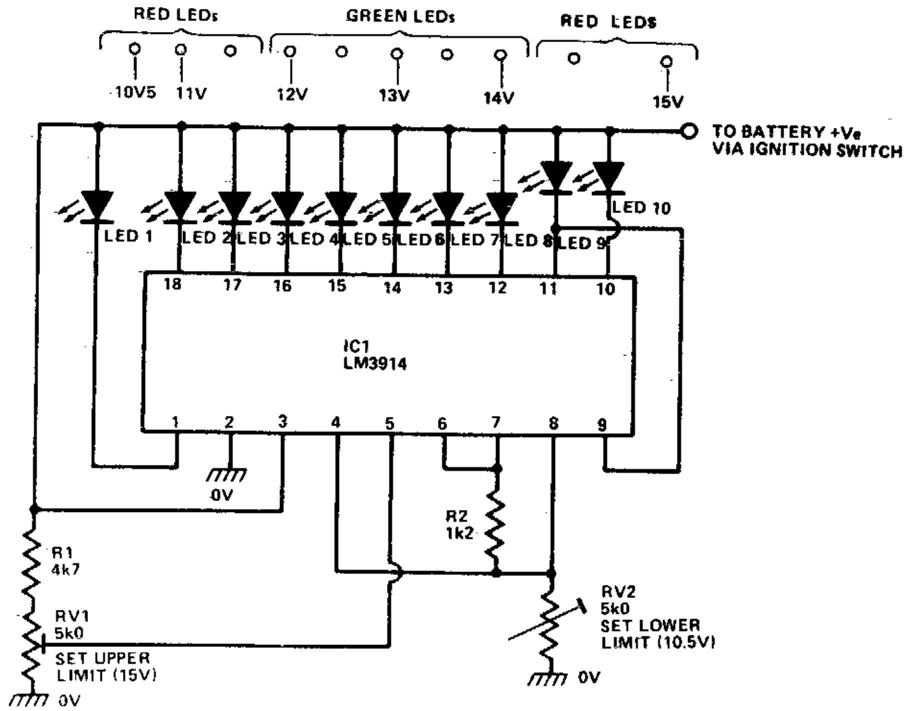


Fig. 10-6

Circuit Notes

The LM3914 acts as a LED-driving voltometer that has its basic maximum and minimum readings determined by the values of R2 and RV2. When correctly adjusted, the unit actually covers the 2.5 volt to 3.6 volt range, but it is made to read a supply voltage span of 10-10.5 volts to 15 volts by interposing potential divider R1-RV1 between the supply line and the pin-5 input terminal of the IC. The IC is configured to give a 'dot' display, in which only one of the ten LEDs is illuminated at any given time. If the supply voltage is below 10.5 volts none of the LEDs illuminate. If the supply equals or exceeds 15 volts, LED 10 illuminates.

TACHOMETER

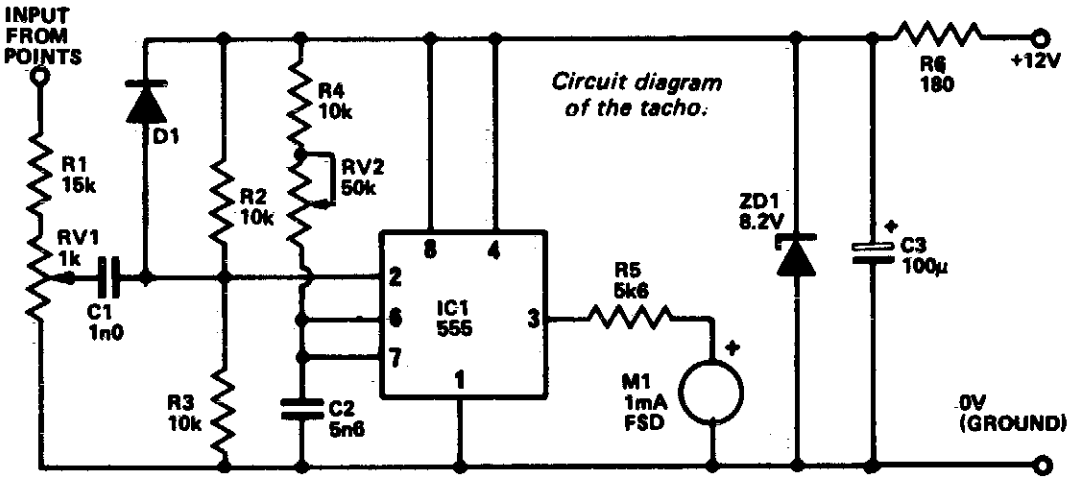


Fig. 10-7

Circuit Notes

An electrical signal taken from the low tension side of the distributor is converted into a voltage proportional to engine rpm and this voltage is displayed on a meter calibrated accordingly. The 555 timer IC is used as a monostable which, in effect, converts the signal pulse from the breaker points to a single positive pulse the width of which is determined by the value of R4 + RV2 and C2. Resistors R2 and R3 set a voltage of about 4 volts at pin 2 of IC1. The IC is triggered if this voltage is reduced to less than approximately 2.7 volts (⅓ of supply voltage), and this occurs due to the voltage swing when the breaker points open. An adjustment potentiometer RV1 enables the input level to be set to avoid false triggering. Zener diode ZD1 and the 180 ohm resistor stabilize the unit against voltage variations.

HIGH SPEED WARNING DEVICE

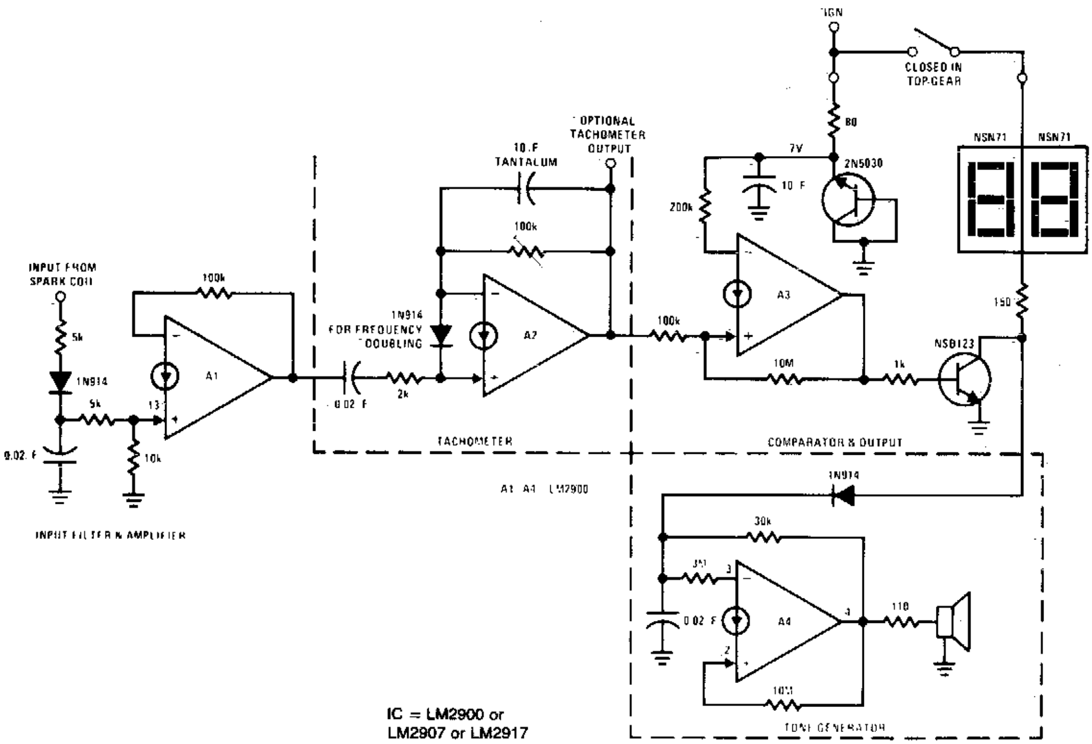


Fig. 10-8

Circuit Notes

A1 amplifies and regulates the signal from the spark coil. A2 converts frequency to voltage so that its output is a voltage proportional to engine rpm. A3 compares the tachometer voltage with the reference voltage and turns on the output transistor at the set speed. Amplifier A4 is used to generate an audible tone whenever the set speed is exceeded.

BREAKER POINT DWELL METER

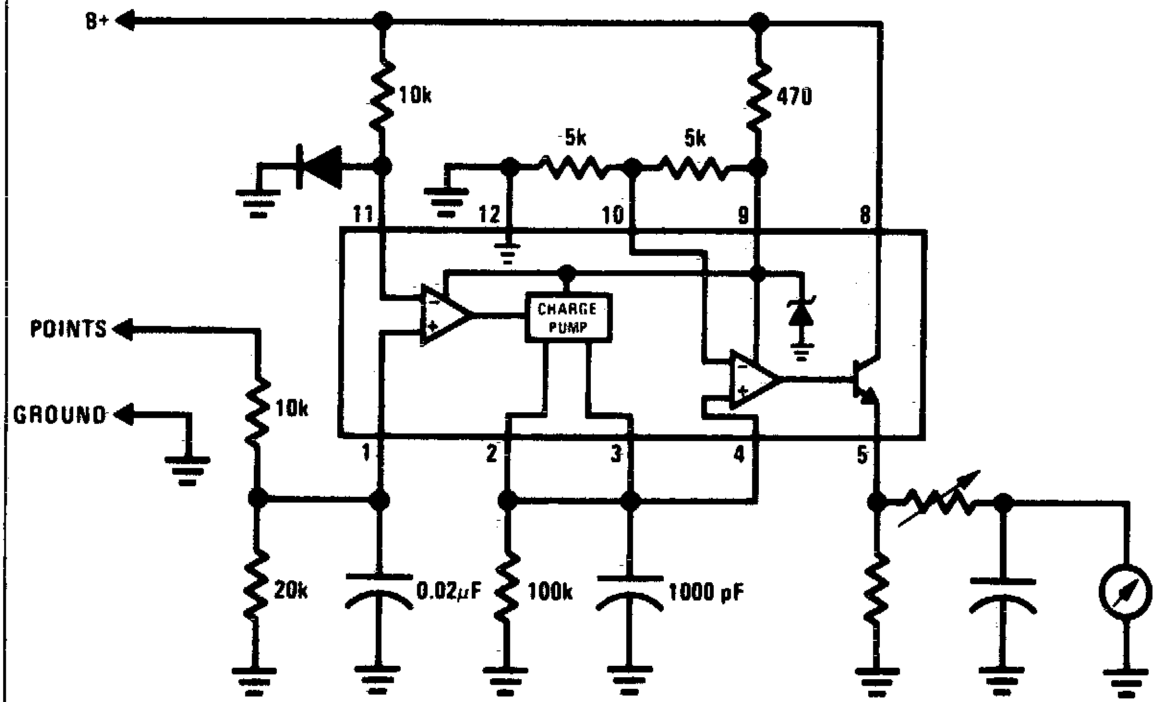
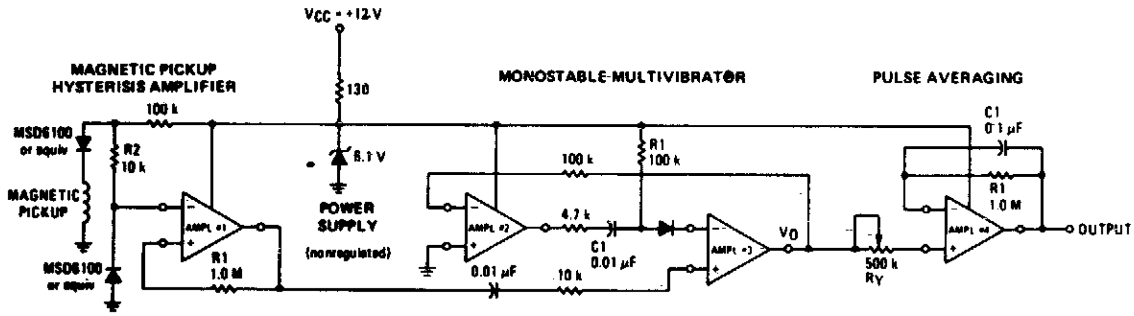


Fig. 10-9

TACHOMETER



Hysterisis Voltage for Switching

$$V_H = \frac{A_I R2}{R1} \quad (V_{CC} - 1.6)$$

Timing-Interval: $t = 0.7\ R1\,C1$

$$V_{p\text{-}p} = \frac{(V_O - 0.6) \cdot A_I \cdot t}{R_Y C1}$$

Fig. 10-10

CAPACITOR DISCHARGE IGNITION SYSTEM

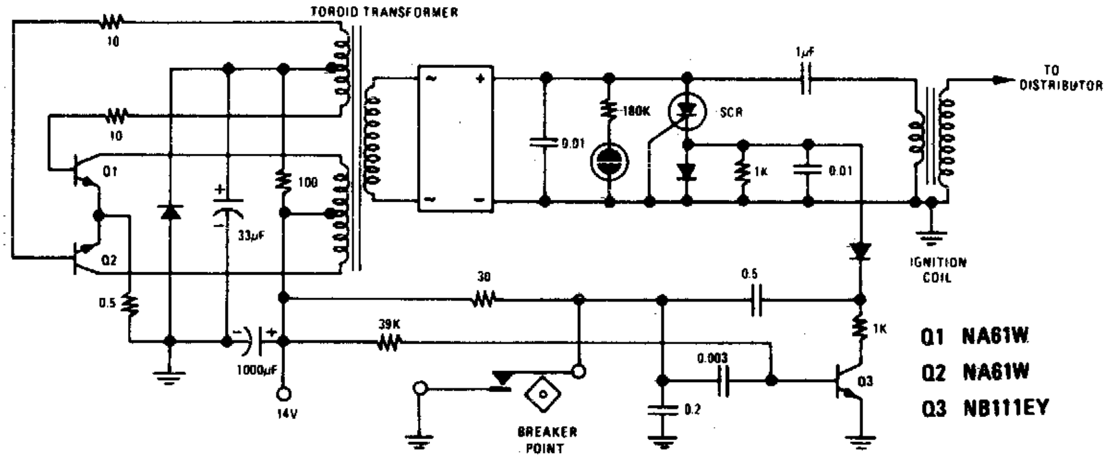


Fig. 10-11

WINDSHIELD WIPER CONTROL

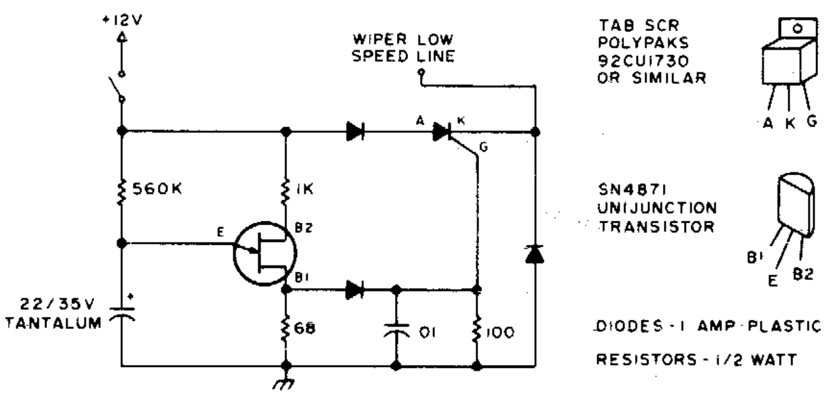


Fig. 10-12

Circuit Notes

Here's a good way to set windshield wipers on an interval circuit. Only two connections to the car's wiper control, plus ground, are required. Variable control can be accomplished by substituting a 500 K pot in series with a 100 K fixed resistor in place of the 560 K.

AUTO BATTERY CURRENT ANALYZER

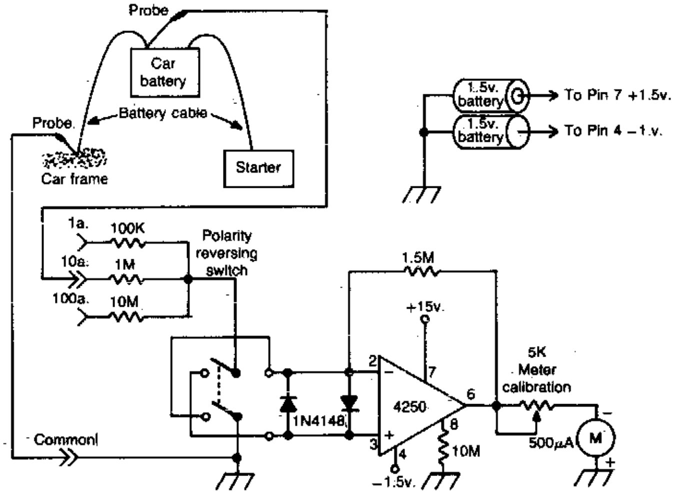
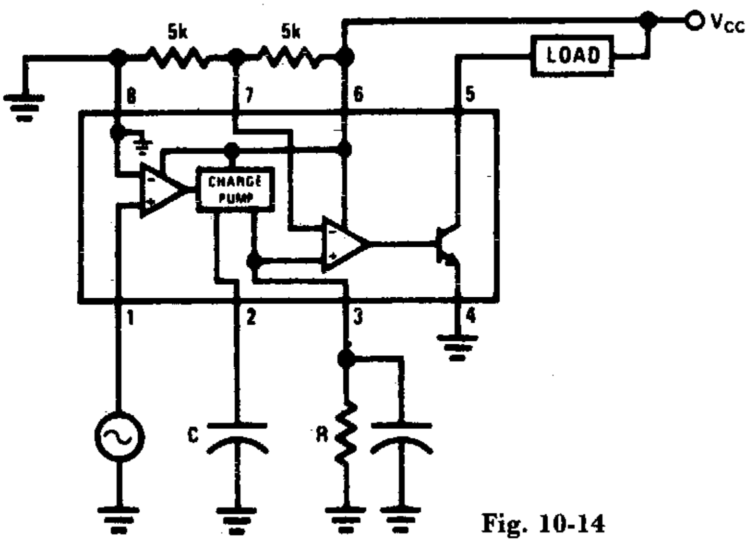


Fig. 10-13

Circuit Notes

This op-amp analyzer can measure the current drawn by any device in a car. The analyzer works by measuring the very small voltage that develops across the battery cables when current flows. To calibrate the unit, measure the current flow somewhere in the car with an accurate ammeter, then adjust the analyzer for that current reading.

SPEED SWITCH



Circuit Notes

Load is energized when

$$f_{in} \geq \frac{1}{2RC}.$$

Fig. 10-14

WINDSHIELD WIPER CONTROLLER

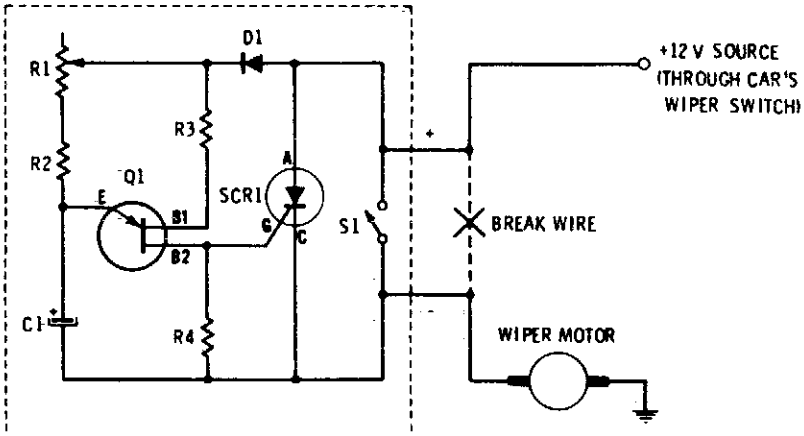


Fig. 10-15

Circuit Notes

This circuit provides complete speed control over car's windshield wipers. They can be slowed down to any rate even down to four sweeps per minute. The controller has two principal circuits: The rate-determining circuit—a unijunction transistor connected as a freerunning oscillator, and the silicon-controlled rectifier which is the actuator.

WINDSHIELD WIPER HESITATION CONTROL UNIT

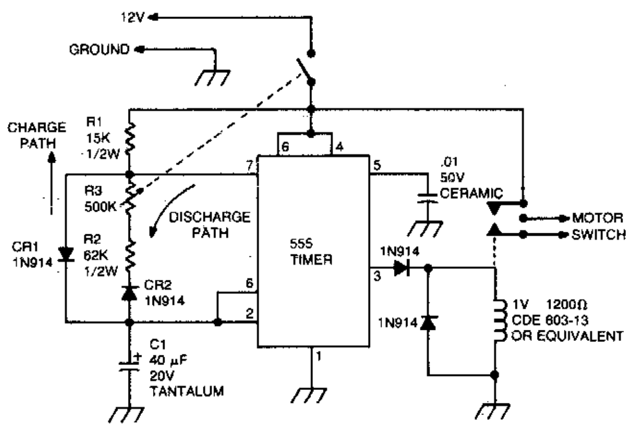


Fig. 10-16

Circuit Notes

This circuit uses the 555 timer in the astable or oscillatory mode. The length of time the timer is off is a function of the values of C1, R2, and R3. The potentiometer which controls the amount of "hesitation". (Approximately 2 to 15 seconds.) R2 provides a minimum time delay when R3 is at its zero ohms position.

ICE WARNING AND LIGHTS REMINDER

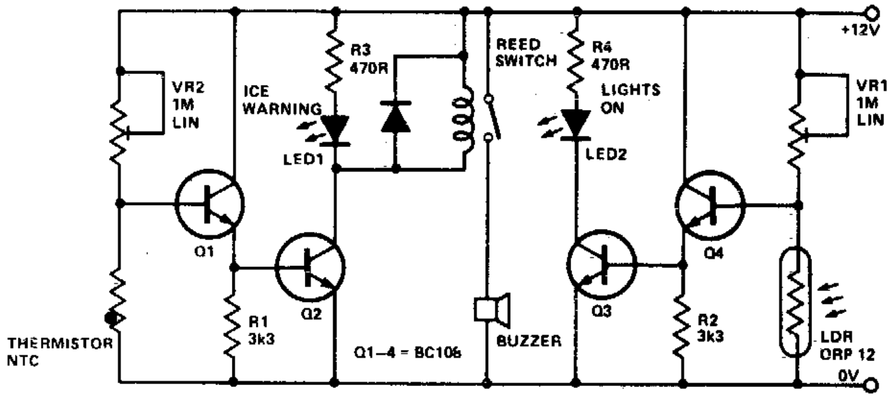
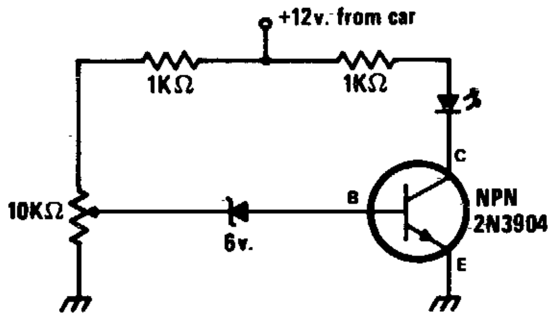


Fig. 10-17

Circuit Notes

This device will tell a driver if his lights should be on and will warn him if the outside temperature is nearing zero by lighting a LED and sounding a buzzer9 VR1 adjusts sensitivity for temperature, VR2 for light. Both thermistor and LDR should be well protected. Most high gain NPN transistors will work.

CAR BATTERY MONITOR



Circuit Notes

Warning light (LED) indicates when battery voltage falls below level set by 10 K pot. Can indicate that battery is defective or needs charging if cranking drops battery voltage below preset "safe" limit.

Fig. 10-18

HEADLIGHT DELAY UNIT

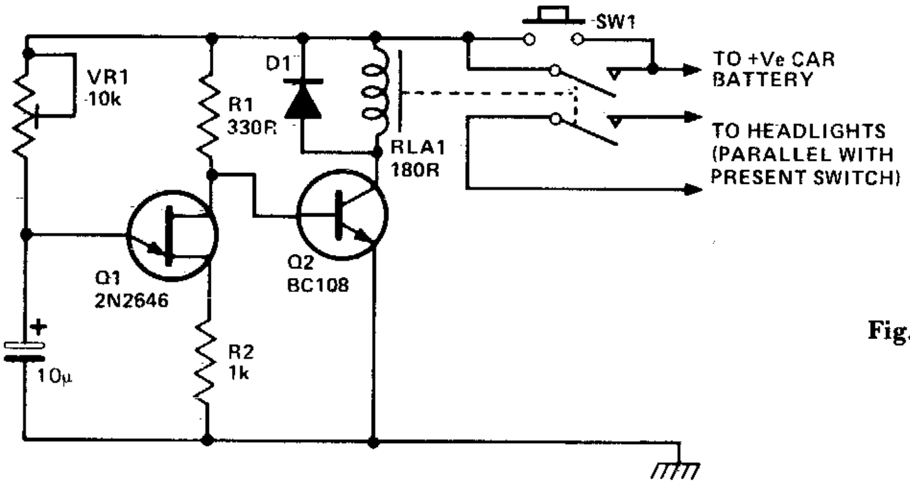


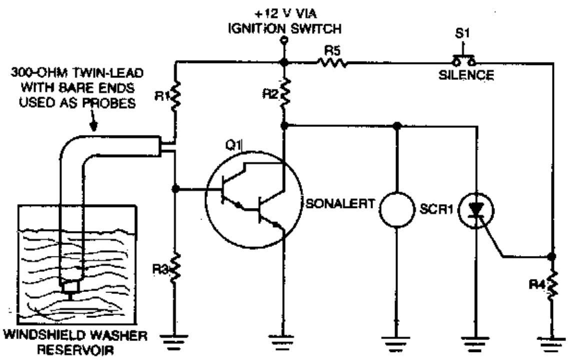
Fig. 10-19

Circuit Notes

This circuit will operate a car's headlights for a predetermined time to light up the driveway or path after the driver has left the car. SQ1 is pushed and Q2 is turned on closing the relay and turning on the car's headlights. C1 begins to charge through VR1 until Q1 turns on, turning Q2 off. The relay will then open switching off both the lights and the unit. The delay is governed by the time taken for the capacitor to charge, which is about one minute.

WINDSHIELD WASHER FLUID WATCHER

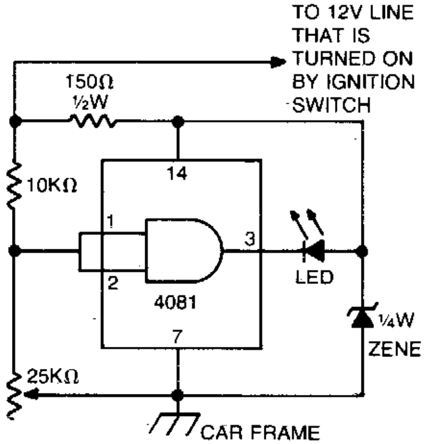
Fig. 10-20



Circuit Notes

This circuit relies upon the minute current between two conductive probes suspended in a washer fluid reservoir. When the level is below the probes, Q1 turns on and the Sonolert sounds.

CAR BATTERY CONDITION CHECKER

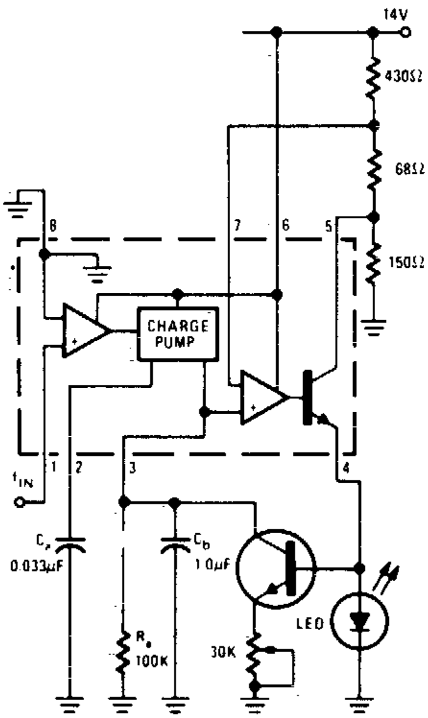


Circuit Notes

This circuit uses an LED and 4081 CMOS integrated circuit. The variable resistor sets the voltage at which the LED turns on. Set the control so that the LED lights when the voltage from the car's ignition switch drops below 13.8 volts. The LED normally will light every now and then for a short period of time. But, if it stays on for very long, your electrical system is in trouble.

Fig. 10-21

OVERSPEED INDICATOR



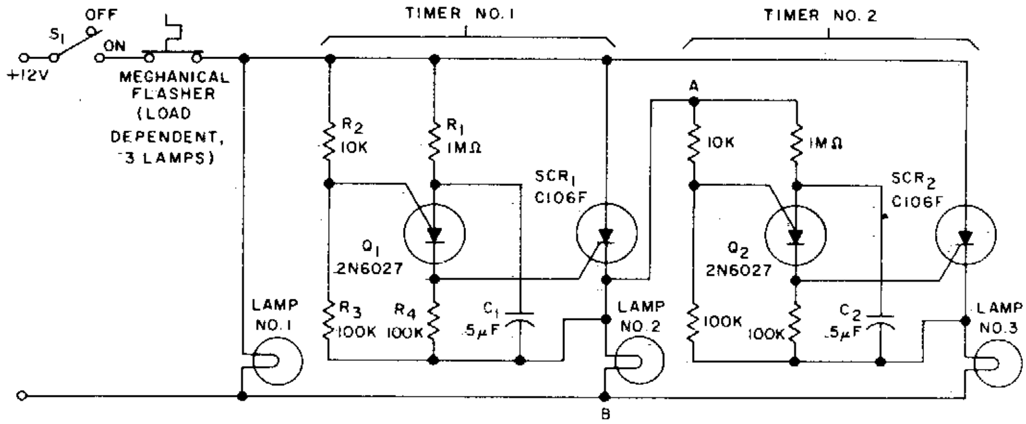
FLASHING BEGINS WHEN f_{IN} > 100 Hz
FLASH RATE INCREASES WITH INPUT FREQUENCY
INCREASE BEYOND TRIP POINT

Circuit Notes

An op-amp comparator is used to compare the converter output with a dc threshold voltage. The circuit flashes the LED when the input frequency exceeds 100 Hz. Increases in frequency raise the average current out of terminal 3 so that frequencies above 100 Hz reduce the charge time of C2, increasing the LED flashing rate. IC = LM2907 or LM2917

Fig. 10-22

SEQUENTIAL FLASHER FOR AUTOMOTIVE TURN SIGNALS



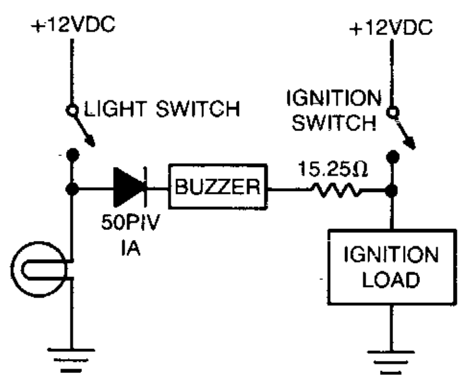
Circuit Notes

When the turn signal switch S1 is closed, lamp #1 will be activated and capacitor C1 will charge to the triggered voltage of Q1. As soon as the anode voltage on Q1 exceeds its gate voltage by 0.5 V, Q1 will switch into the low resistance mode, thereby triggering SCR1 to activate lamp #2 and the second timing circuit. After Q2 switches into the low resistance state, SCR2 will be triggered to activate lamp #3. When the thermal flasher interrupts the current to all three lamps, SCR1 and SCR2 are commutated and the circuit is ready for another cycle.

Fig. 10-23

AUTO LIGHTS-ON REMINDER

Fig. 10-24



Circuit Notes

The alarm is composed of a diode, buzzer, and limiting resistor. The diode serves as a switch which allows the buzzer to sound off only when the light switch is closed and the ignition is turned off.

11

Battery Chargers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

12 V BATTERY CHARGER

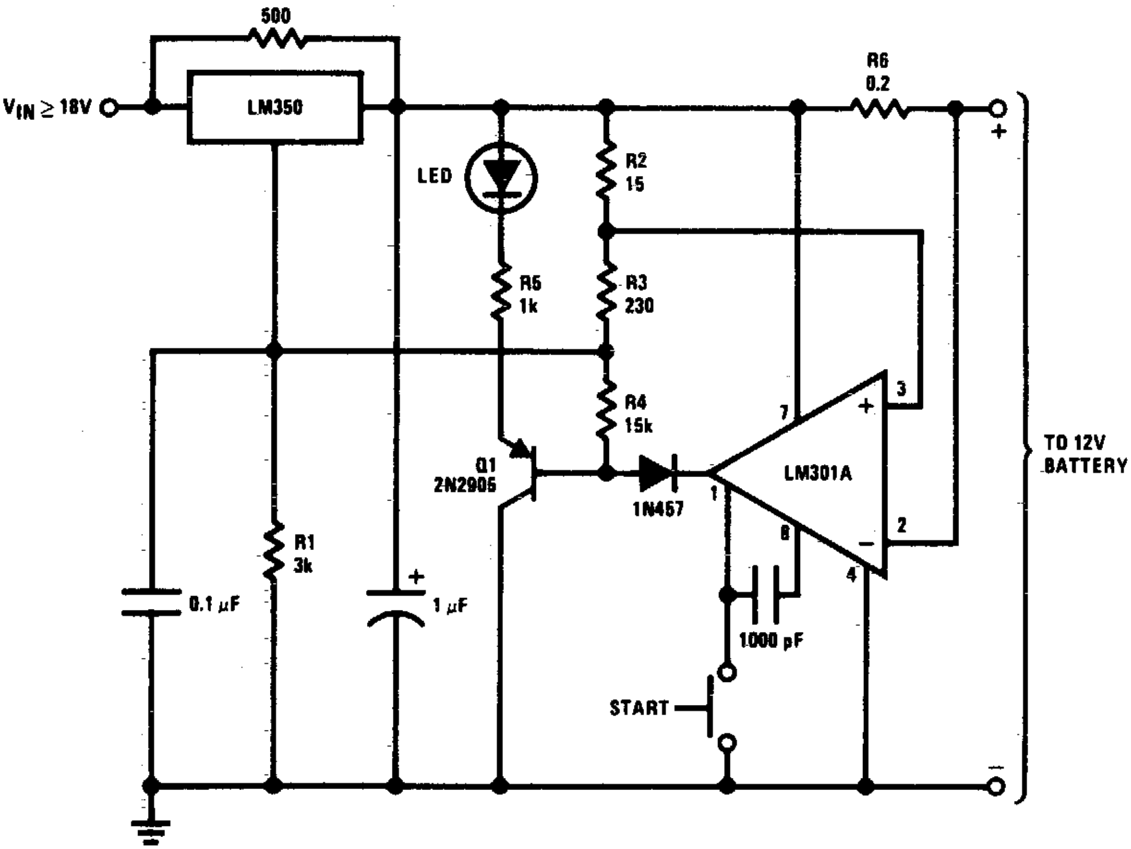


Fig. 11-1

Circuit Notes

This circuit is a high performance charger for gelled electrolyte lead-acid batteries. Charger quickly recharges battery and shuts off at full charge. Initially, charging current is limited to 2A. As the battery voltage rises, current to the battery decreases, and when the current has decreased to 150 mA, the charger switches to a lower float voltage preventing overcharge. When the start switch is pushed, the output of the charger goes to 14.5 V. As the battery approaches full charge, the charging current decreases and the output voltage is reduced from 14.5 V to about 12.5 V terminating the charging. Transistor Q1 then lights the LED as a visual indication of full charge.

SIMPLE NI-CAD BATTERY CHARGER

PARTS LIST FOR
NICAD BATTERY CHARGER

C1—100-μF, 50-V electrolytic capacitor
D1—1-A, 400 PIV-silicon rectifier
Q1—40-W, pnp power transistor
R1—2000-ohm potentiometer
T1—24-Vac, 117-Vac primary filament transformer

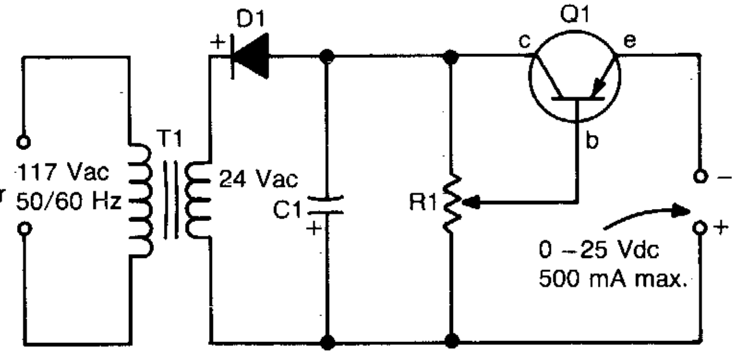
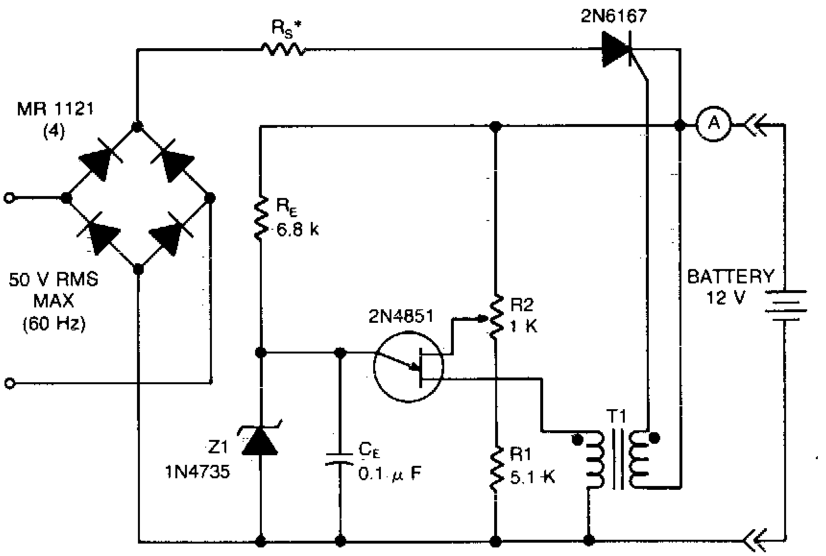


Fig. 11-2

Circuit Notes

This circuit provides an adjustable output voltage up to 35 Vdc and maximum output current of 50 mA. Transistor Q1 dissipates quite a bit of heat and must be mounted on a heatsink.

12 V BATTERY CHARGER CONTROL (20 AMPS RMS MAX.)

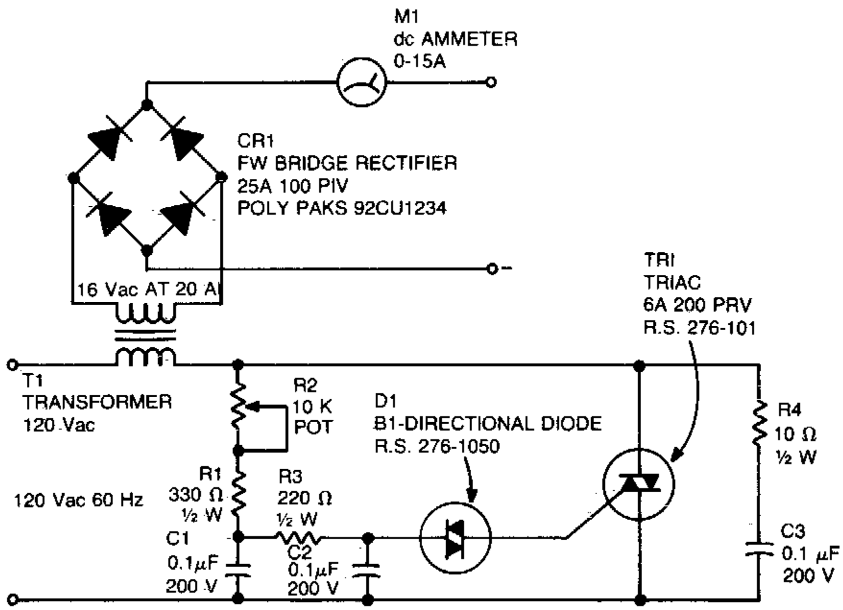


T1 - PRIMARY = 30 TURNS #22
SECONDARY = 45 TURNS #22
CORE = FERROXCUBE 203 F 181-3C3
* R_S - SERIES RESISTANCE TO LIMIT CURRENT THROUGH SCR.
2N6167 IS RATED AT 20 AMPS RMS.

Fig. 11-3

BATTERY CHARGER

Fig. 11-4



Circuit Notes

A diac is used in the gate circuit to provide a threshold level for firing the triac. C3 and R4 provide a transient suppression network. R1, R2, R3, C1, and C2 provide a phase-shift network for the signal being applied to the gate. R1 is selected to limit the maximum charging current at full rotation of R2.

AUTOMATIC SHUTOFF BATTERY CHARGER

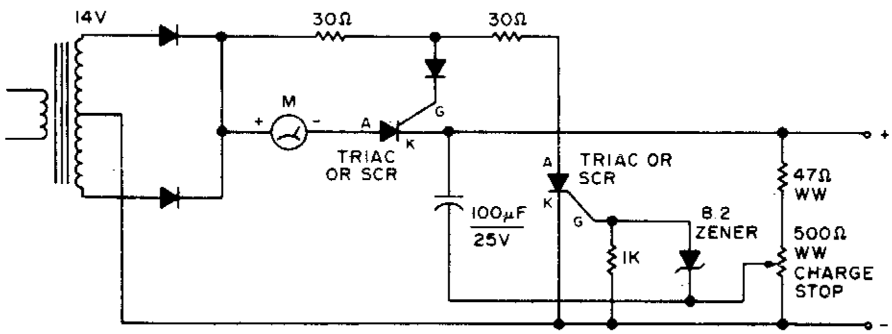


Fig. 11-5

Circuit Notes

Adjust by setting the 500 ohm resistor while attached to a fully charged battery.

200 mA-HOUR, 12 V NI-CAD BATTERY CHARGER

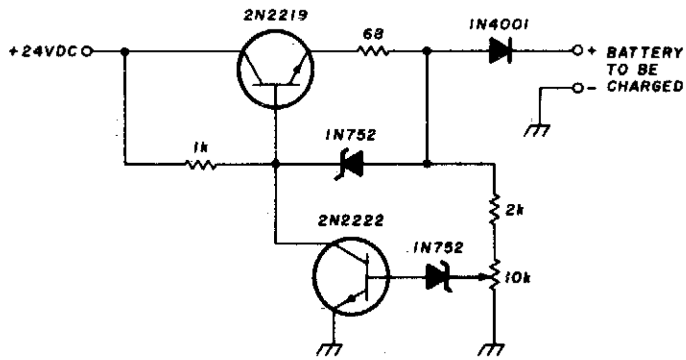


Fig. 11-6

Circuit Notes

This circuit charges the battery at 75 mA until the battery is charged, then it reduces the current to a trickle rate. It will completely recharge a dead battery in four hours and the battery can be left in the charger indefinitely. To set the shut-off point, connect a 270-ohm, 2-watt resistor across the charge terminals and adjust the pot for 15.5 volts across the resistor.

NI-CAD CHARGER WITH CURRENT AND VOLTAGE LIMITING

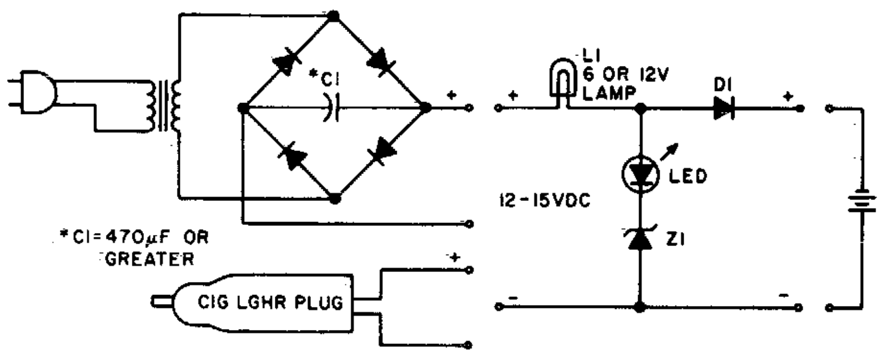


Fig. 11-7

Circuit Notes

Lamp L1 will glow brightly and the LED will be out when the battery is low and being charged, but the LED will be bright and the light bulb dim when the battery is almost ready. L1 should be a light bulb rated for the current you want (usually the battery capacity divided by 10). Diode D1 should be at least 1 A, and Z1 is a 1 W zener diode with a voltage determined by the full-charge battery voltage minus 1.5 V. After the battery is fully charged, the circuit will float it at about battery capacity divided by 100 mA.

AUTOMOTIVE CHARGER FOR NI-CAD BATTERY PACKS

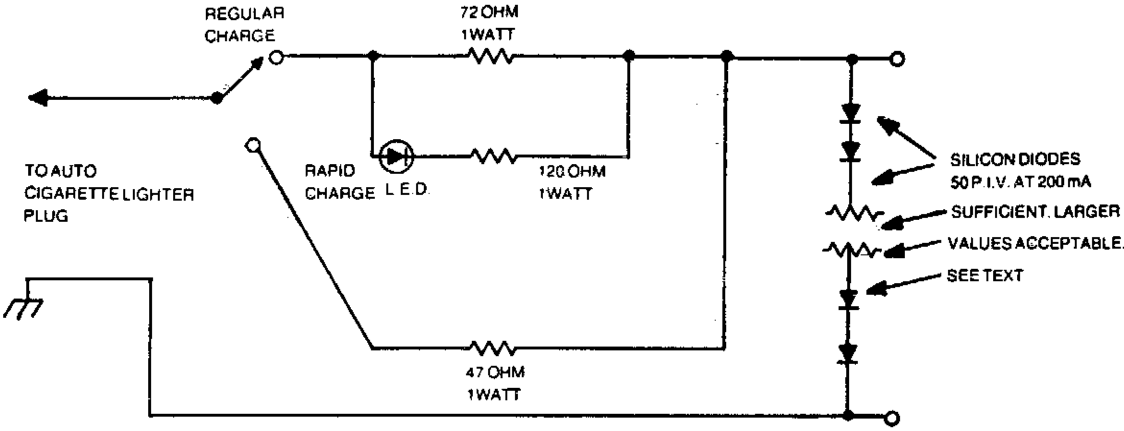


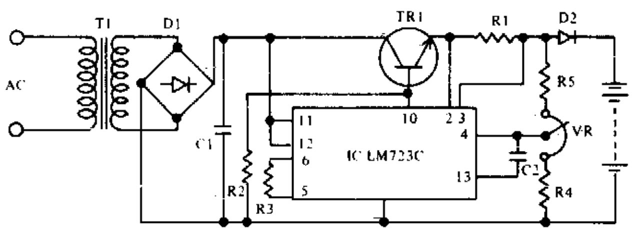
Fig. 11-8

Circuit Notes

The number of silicon diodes across the output is determined by the voltage of the battery pack. Figure each diode at 0.7 volt. For example, a 10.9- volt pack would require 10.9/0.7 = 15.57, or 16 diodes.

CONSTANT-VOLTAGE, CURRENT-LIMITED CHARGER

IC LM723C VOLTAGE REGULATOR (FOR 12V dc OUTPUT 0.42A MAX.)



T1 TRANSFORMER, DC 13V (RMS), 1-3A (RMS)
D1, D2 100V 1A DIODE
C1 50V, 470μF ELECTROLYTIC CONDENSER
TR1 MJ2840 10A 60V 150W (MOTOROLA)
IC LM723C (NATIONAL SEMICONDUCTOR)
R1 4.7 OHM 1/2W 3P
R2 5.1K OHM 1/4W
R3 3.9K OHM 1/4W
R4 7.5K OHM 1/4W
R5 8.2K OHM 1/4W
VR 2K OHM
C2 50V 1000PF

Circuit Notes

For 12 V sealed lead-acid batteries.

Fig. 11-9

NI-CAD CHARGER

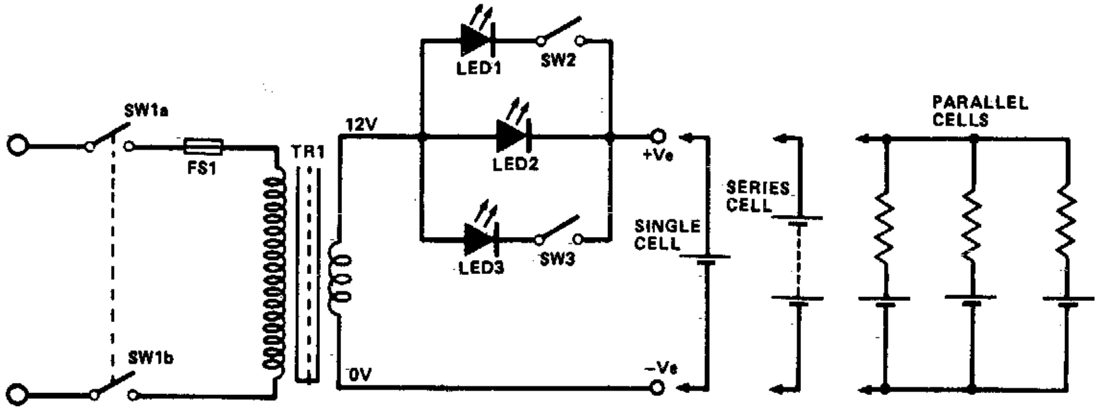


Fig. 11-10

Circuit Notes

This circuit uses constant current LEDs to adjust charging current. It makes use of LEDs that pass a constant current of about 15 mA for an applied voltage range of 2-18 V. They can be paralleled to give any multiple of 15 mA and they light up when current is flowing. The circuit will charge a single cell at 15, 30 or 45 mA or cells in series up to the rated supply voltage limit (about 14 V).

SIMPLE NI-CAD BATTERY ZAPPER

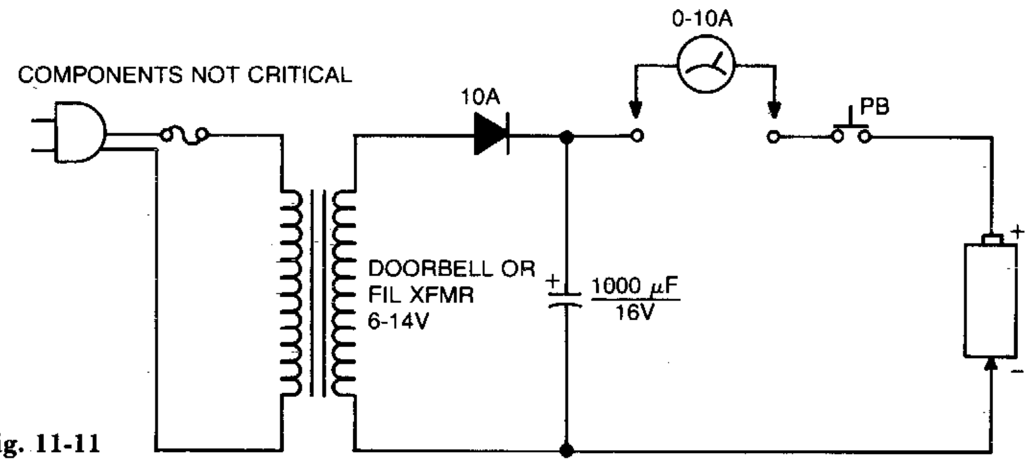


Fig. 11-11

Circuit Notes

This circuit is used to clear internal shorts in nickel cadmium batteries. To operate, connect ni-cad to output and press the pushbutton for three seconds.

BATTERY CHARGING REGULATOR

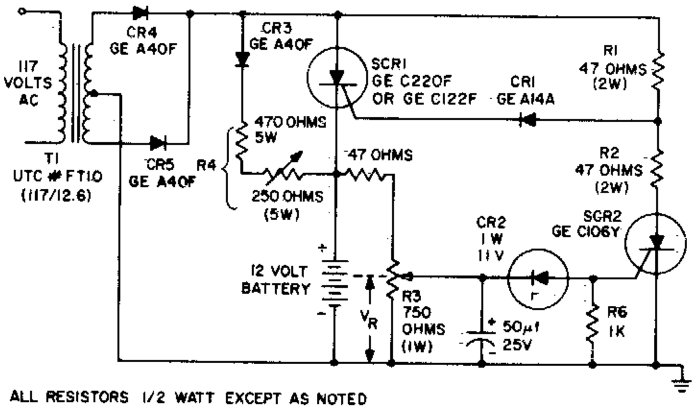


Fig. 11-12

Circuit Notes

The circuit is capable of charging a 12-volt battery at up to a six ampere rate. Other voltages and currents, from 6 to 600 volts and up to 300 amperes, can be accommodated by suitable component selection. When the battery voltage reaches its fully charged level, the charging SCR shuts off, and a trickle charge as determined by the value of R4 continues to flow.

LOW-COST TRICKLE CHARGER FOR 12 V STORAGE BATTERY

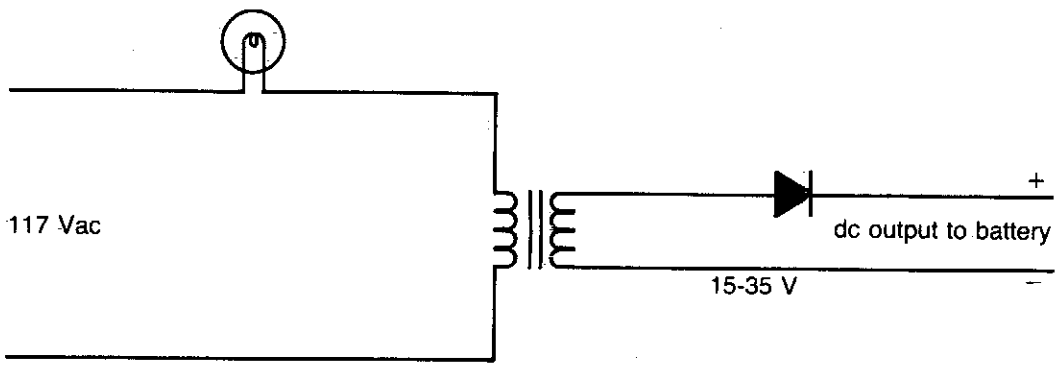
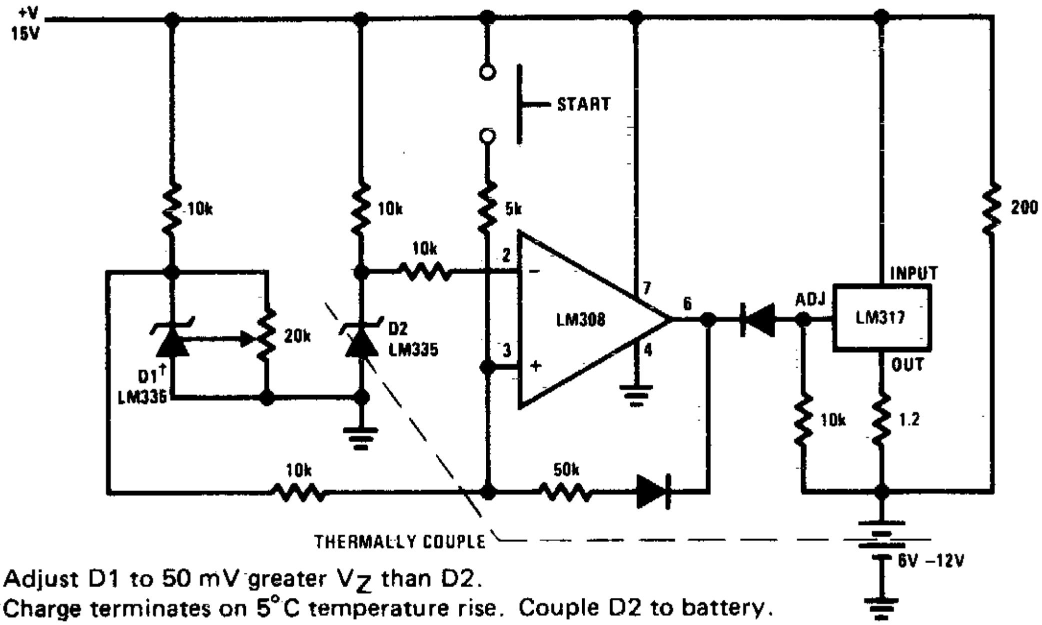


Fig. 11-13

Circuit Notes

Charge rate can be varied and is based on the size of bulb.

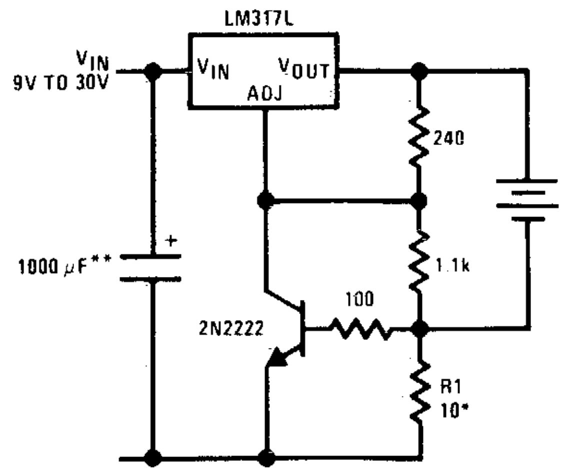
FAST CHARGER FOR NI-CAD BATTERIES



Adjust D1 to 50 mV greater V_Z than D2.
Charge terminates on 5°C temperature rise. Couple D2 to battery.

Fig. 11-14

CURRENT LIMITED 6 V CHARGER



* Sets peak current, $I_{PEAK} = 0.6V/R1$
** 1000 μF is recommended to filter out any input transients.

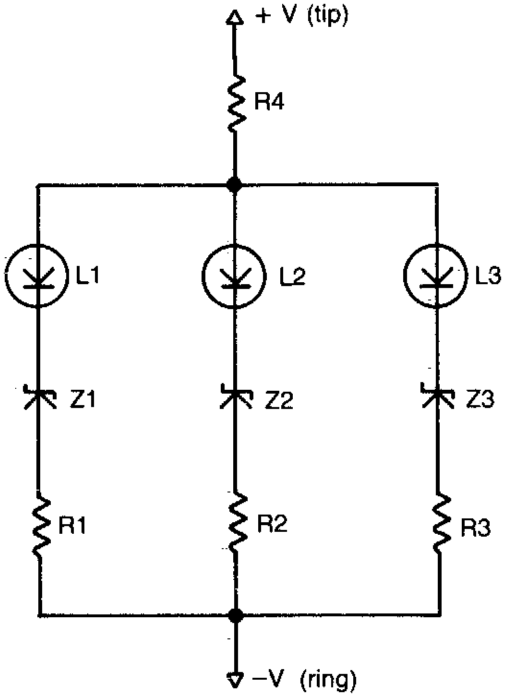
Fig. 11-15

12

Battery Monitors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SOLID-STATE BATTERY VOLTAGE INDICATOR

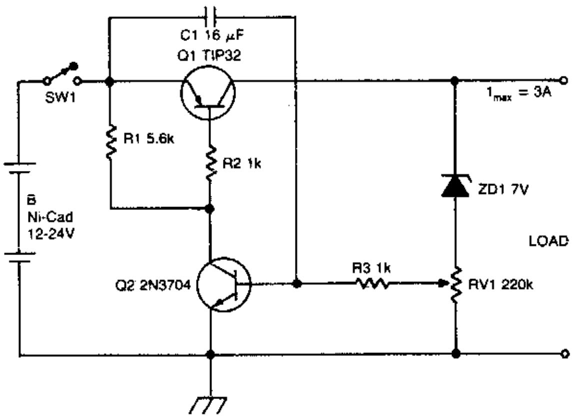


R1, R2, R3 = 47 Ω
R4 = 39 Ω
Z1 = 9.8 volt zener diode
Z2 = 11.1 volt zener diode
Z3 = 11.5 volt zener diode
L1 – L3 = light emitting diodes

Two lights on - OK (L1 + L2)
One light on - low voltage (L1 only)
Three lights on - overvoltage (L1 + L2 + L3)

Fig. 12-1

NI-CAD DISCHARGE LIMITER



Circuit Notes

The circuit disconnects the battery from the load when output voltage falls below a preset level. C1 charges through R1 and turns on Q2. Collector current flows through R2 turning Q1 on and battery is connected to the load. When the output voltage falls below a point set by RV1, Q2 turns off, Q1 turns off and further discharge of the battery is prevented.

Fig. 12-2

BATTERY CONDITION INDICATOR

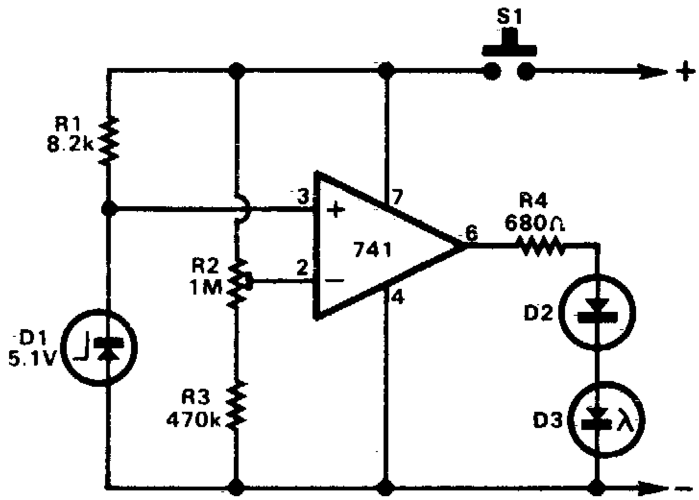


Fig. 12-3

Circuit Notes

A 741 op amp is employed as a voltage comparator. The noninverting input is connected to zener reference source. Reference voltage is 5.1V. R2 is adjusted so that the voltage at the inverting input is half the supply voltage. When supply is higher than 10.2V, the LED will not light. When the supply falls just fractionally below the 10.2V level, the IC inverting input will be slightly negative of the noninverting input, and the output will swing fully positive. The LED will light, indicating that the supply voltage has fallen to the preset threshold level. The LED can be made to light at other voltages by adjusting R2.

EQUIPMENT ON REMINDER

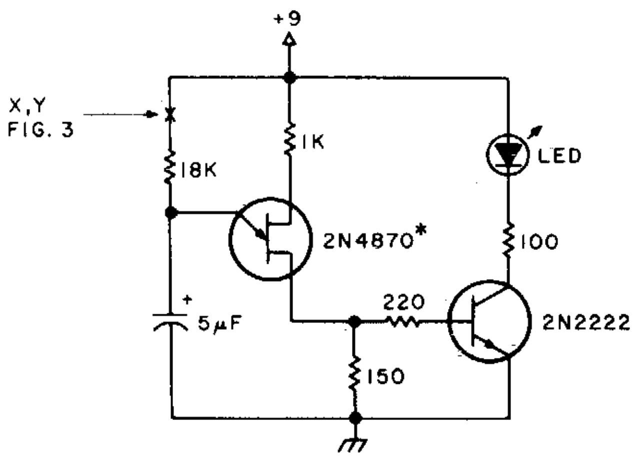


Fig. 12-4

*RADIO SHACK
RS 276-2029
OR ANY TYPE UJT

Circuit Notes

Due to the low duty cycle of flashing LED, the average current drain is 1 mA or less.

BATTERY CHARGE/DISCHARGE INDICATOR

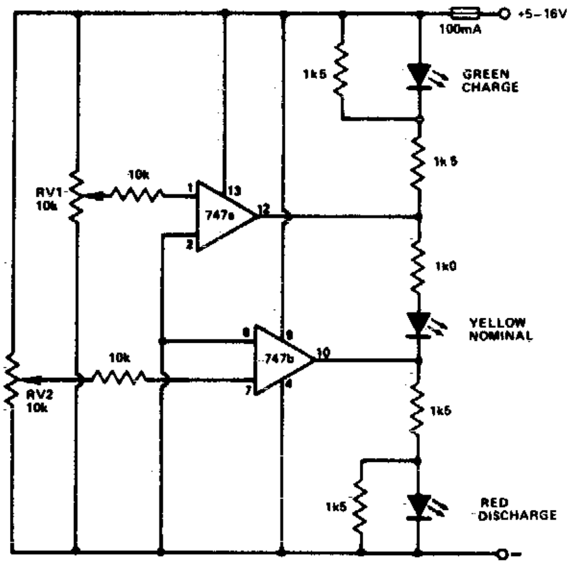


Fig. 12-5

Circuit Notes

This circuit monitors car battery voltage. It provides an indication of nominal supply voltage as well as low or high voltage. RV1 and RV2 adjust the point at which the red/yellow and yellow/green LEDs are on or off. For example the red LED comes on at 11V, and the green LED at 12V. The yellow LED is on between these values.

PRECISION BATTERY VOLTAGE MONITOR FOR HTS

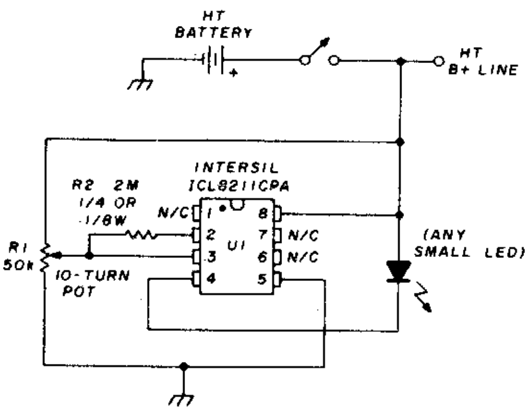
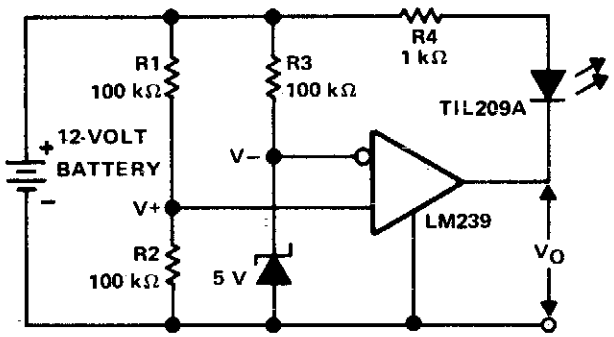


Fig. 12-6

Circuit Notes

The precision voltage-monitor chip contains a temperature-compensated voltage reference. R1 divides down the battery voltage to match the built-in reference voltage of IC1 (1.15 volts). When the voltage at pin 3 falls below 1.15 volts, pin 4 supplies a constant current of 7 mA to drive a small LED. About 0.2 volt of hysteresis is added with R2. Without hysteresis, the LED could flicker on and off when the monitored voltage varies around the set point, as might be the case on voice peaks during receive.

LOW-VOLTAGE MONITOR



a. SCHEMATIC OF CIRCUIT FOR LOW-VOLTAGE INDICATOR

Fig. 12-7

Circuit Notes

This circuit monitors the voltage of a battery and warns the operator when the battery voltage is below a preset level by turning on an LED. The values are set for a 12V automobile battery. The preset value is 10 volts.

UNDERVOLTAGE INDICATOR FOR BATTERY OPERATED EQUIPMENT

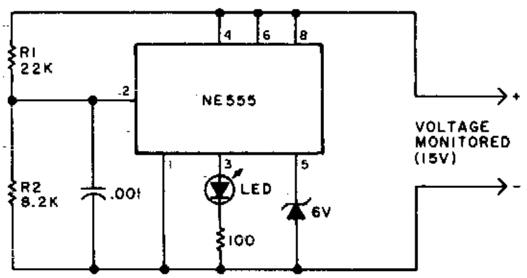


Fig. 12-8

Circuit Notes

Due to the low duty cycle of flashing LED, the average current drain is 1 mA or less. The NE555 will trigger the LED on when the monitored voltage falls to 12 volts. The ratio of R1 to R2 only needs to be changed if it is desired to change the voltage point at which the LED is triggered.

LOW BATTERY INDICATOR

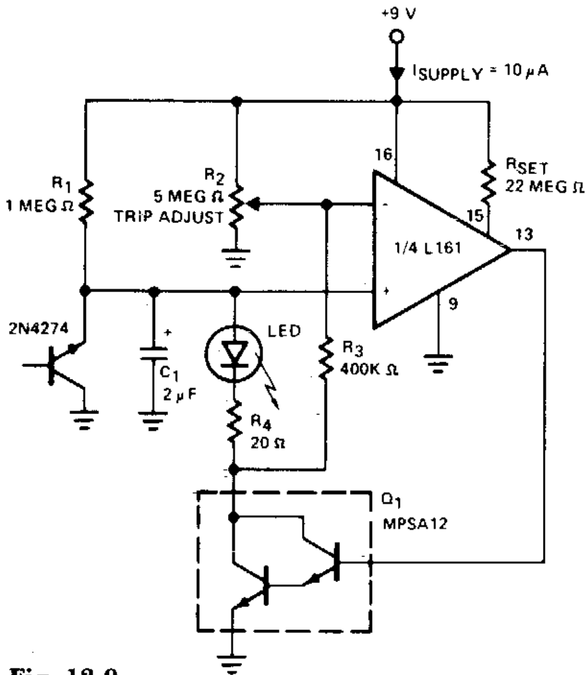


Fig. 12-9

Circuit Notes

The indicator flashes an LED when the battery voltage drops below a certain threshold. 2N4274 emitter-base junction serves as a zener which establishes about 6V on the L161's positive input. As the battery drops, the L161 output goes high. This turns on the Darlington, which discharges C1 through the LED. The interval between flashes is roughly two seconds and gives a low battery warning with only 10 μA average power drain.

BATTERY-LEVEL INDICATOR

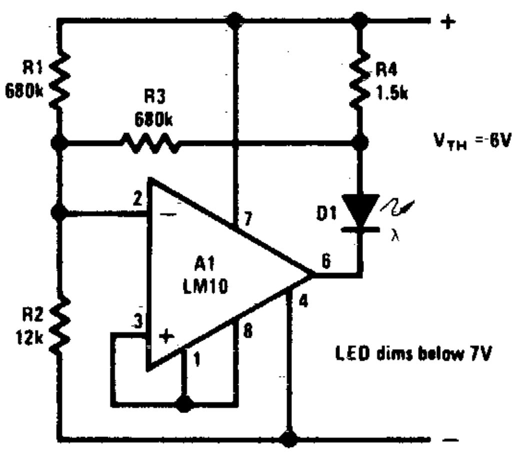


Fig. 12-10

BATTERY-THRESHOLD INDICATOR

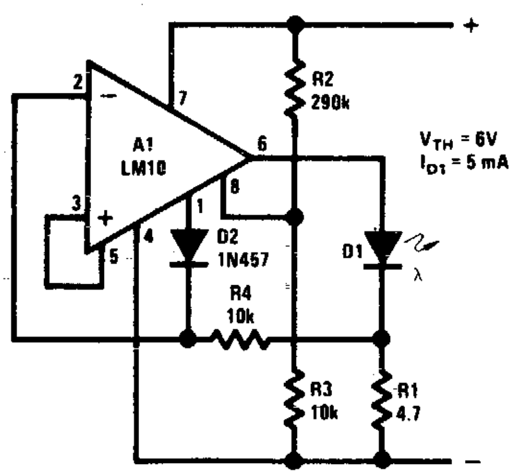


Fig. 12-11

13

Buffers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SINE WAVE OUTPUT BUFFER AMPLIFIER

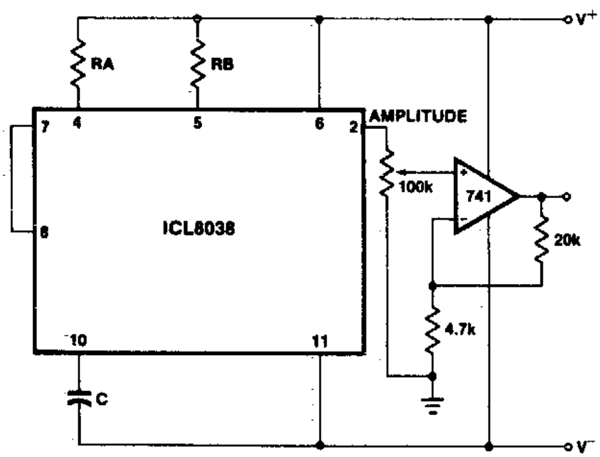


Fig. 13-1

Circuit Notes

The sine wave output has a relatively high output impedance (1K typ). The circuit provides buffering, gain, and amplitude adjustment. A simple op amp follower could also be used.

SINGLE SUPPLY AC BUFFER AMPLIFIER

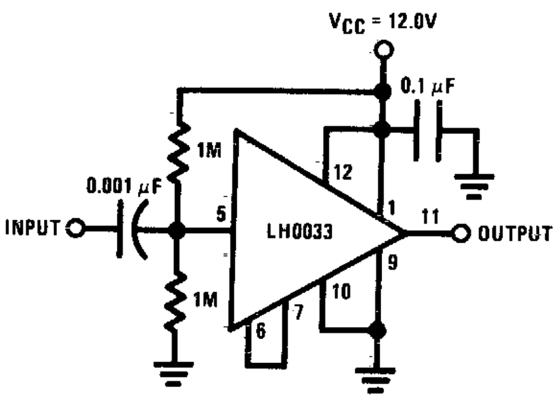


Fig. 13-2

Circuit Notes

The input is dc biased to mid-operating point and is ac coupled. Its input impedance is approximately 500K at low frequencies. For dc loads referenced to ground, the quiescent current is increased by the load current set at the input dc bias voltage.

SINGLE SUPPLY AC BUFFER (HIGH SPEED)

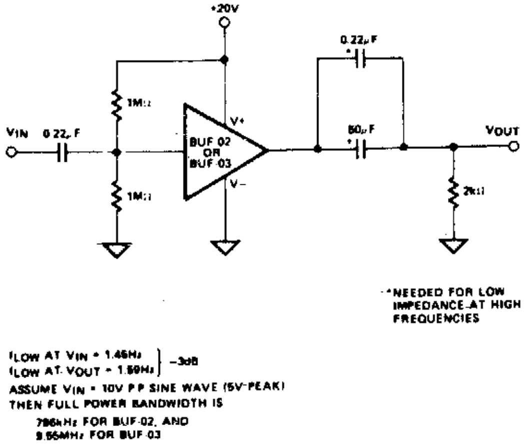


Fig. 13-3

HIGH IMPEDANCE LOW CAPACITANCE WIDEBAND BUFFER

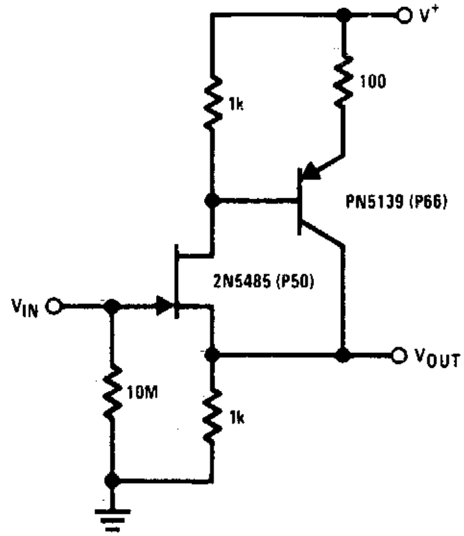


Fig. 13-5

Circuit Notes

The 2N5485 has low input capacitance which makes this compound series-feedback buffer a wide-band unity gain amplifier.

HIGH SPEED 6-BIT A/D BUFFER

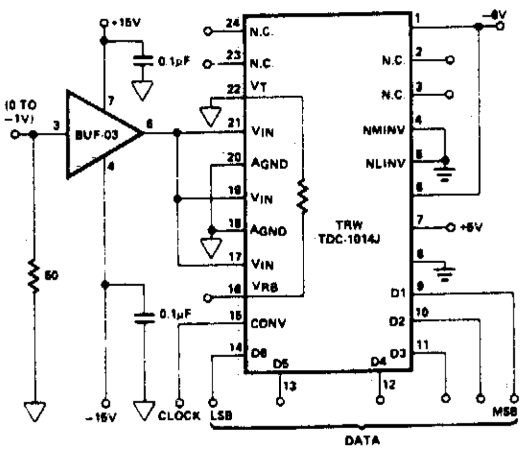


Fig. 13-4

HIGH RESOLUTION ADC INPUT BUFFER

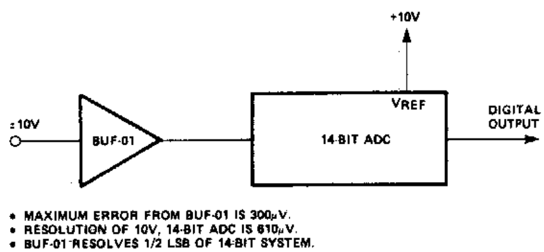


Fig. 13-6

100 × BUFFER AMPLIFIER

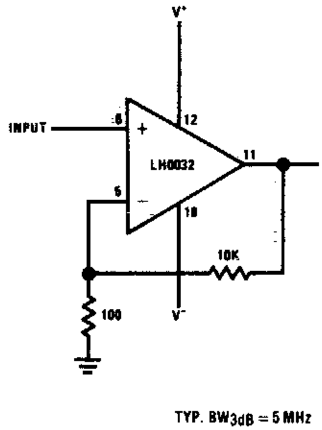


Fig. 13-7

STABLE, HIGH IMPEDANCE BUFFER

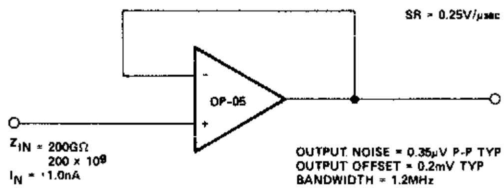


Fig. 13-9

10 × BUFFER AMPLIFIER

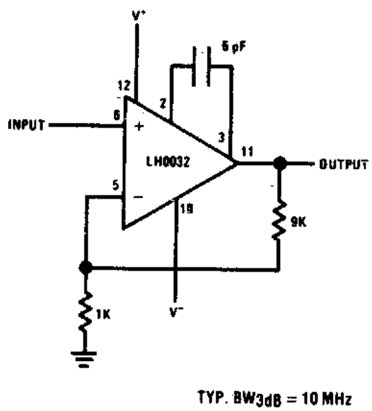
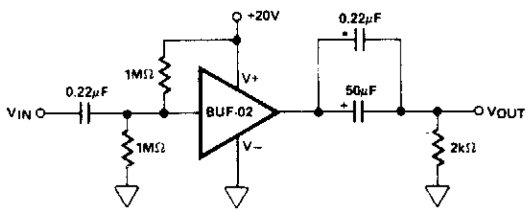


Fig. 13-8

HIGH-SPEED SINGLE-SUPPLY AC BUFFER



f_{LOW} AT V_{IN} = 1.45Hz
f_{LOW} AT V_{OUT} = 1.59Hz −3dB
ASSUME V_{IN} = 10V P-P SINE WAVE (5V PEAK)
THEN FULL POWER BANDWIDTH IS APPROXIMATELY 800kHz.

Fig. 13-10

14

Capacitance (Touch) Operated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Capacitance Relay
- Capacitance Operated, Battery Powered Light
- Touch Sensitive Switch
- Low Current Touch Switch
- Capacitance Switched Light
- Momentary Operation Touch Switch
- Touch Triggered Bistable
- Capacitance Operated Alarm to Foil Purse Snatchers
- Self-Biased Proximity Sensor Works on Detected Changing Fields
- Touch Switch or Proximity Detector
- Finger Touch Touch or Control Switch
- Proximity Detector
- Touch Circuit
- CMOS Touch Switch
- Latching Double-Button Touch Switch

CAPACITANCE RELAY

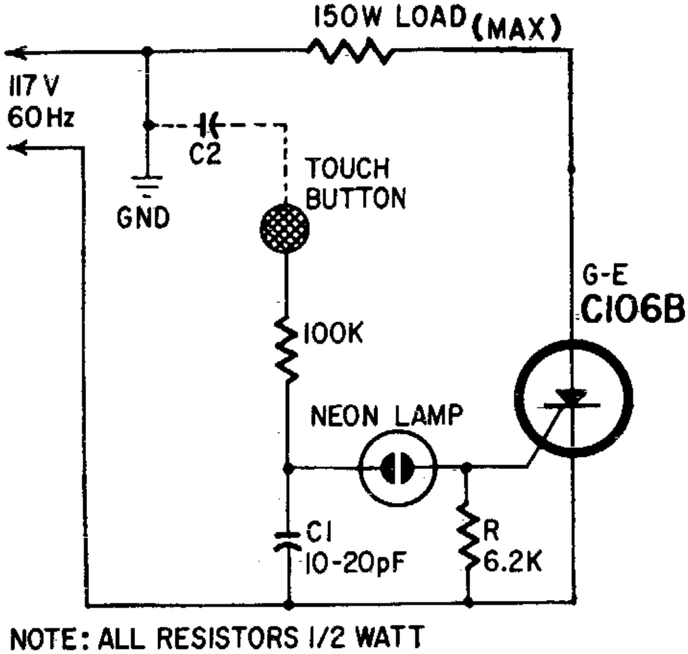


Fig. 14-1

Circuit Notes

Capacitor C1 and body capacitance (C2) of the operator form the voltage divider from the hot side of the ac line to ground. The voltage across C1 is determined by the ratio of C1 to C2. The higher voltage is developed across the smaller capacitor. When no one is close to the touch button, C2 is smaller than C1. When a hand is brought close to the button, C2 is many times larger than C1 and the major portion of the line voltage appears across C1. This voltage fires the neon lamp, C1 and C2 discharge through the SCR gate, causing it to trigger and pass current through the load. The sensitivity of the circuit depends on the area of the touch plate. When the area is large enough, the circuit responds to the proximity of an object rather than to touch. C1 may be made variable so sensitivity can be adjusted.

CAPACITANCE OPERATED, BATTERY POWERED LIGHT

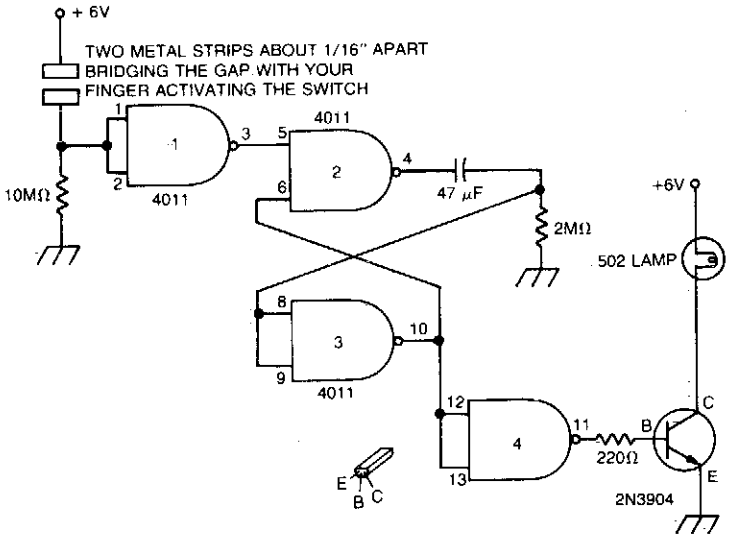


Fig. 14-2

Circuit Notes

Touch the plate and the light will go on and remain on for a time determined by the time constant of the 47 μF capacitor and the 2M resistor.

TOUCH-SENSITIVE SWITCH

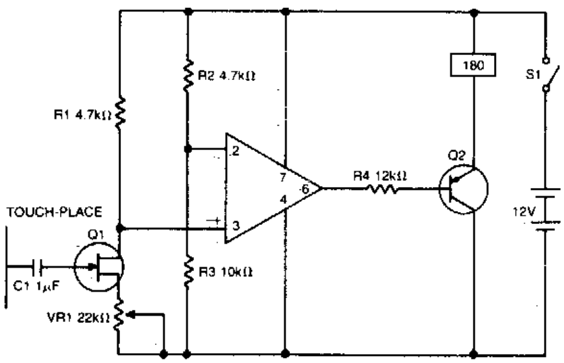


Fig. 14-3

Circuit Notes

A high impedance input is provided by Q1, a general purpose field effect transistor. 741 op amp is used as a sensitive voltage level switch which in turn operates the current Q2, a medium current PNP bipolar transistor, thereby energizing the relay which can be used to control equipment, alarms, etc.

LOW CURRENT TOUCH SWITCH

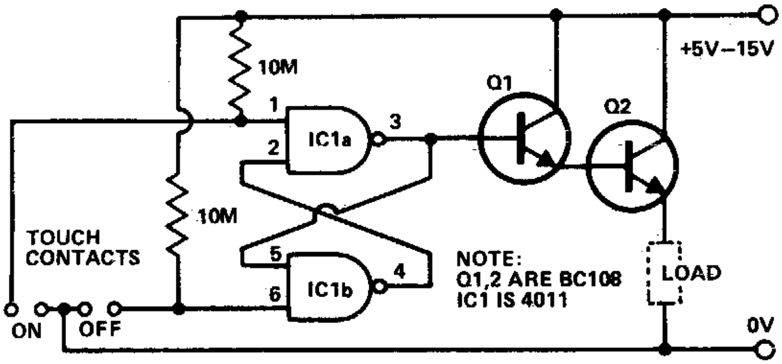


Fig. 14-4

Circuit Notes

Touching the on contacts with a finger brings pin 3 high, turning on the Darlington pair and supplying power to the load (transistor radio etc). Q1 must be a high gain transistor, and Q2 is chosen for the current required by the load circuit.

CAPACITANCE SWITCHED LIGHT

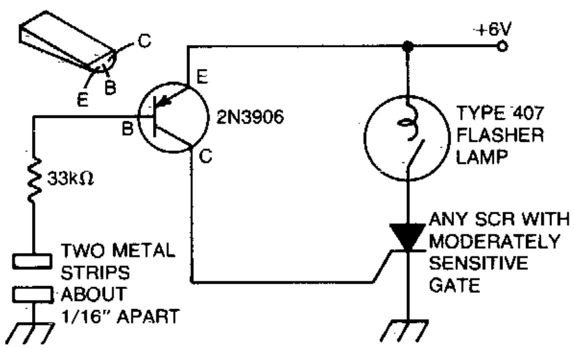


Fig. 14-5

Circuit Notes

The battery-powered light turns on easily, stays on for just a few seconds, and then turns off again. The circuit is triggered when you place a finger across the gap between two strips of metal, about 1/16th inch apart. Enough current will flow through your finger to trigger the SCR after being amplified by the 2N3906. Once the SCR is fired, current will flow through the bulb until its internal bimetal switch turns it off. Once that happens, the SCR will return to its nonconducting state.

MOMENTARY OPERATION TOUCH SWITCH

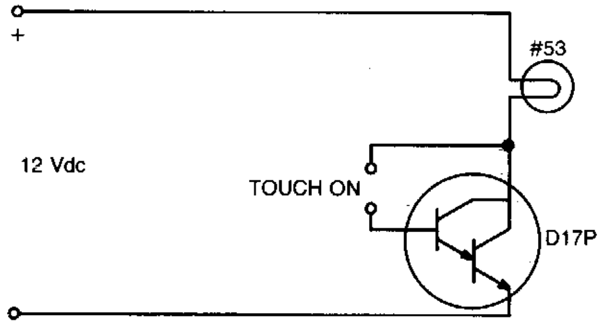


Fig. 14-6

TOUCH TRIGGERED BISTABLE

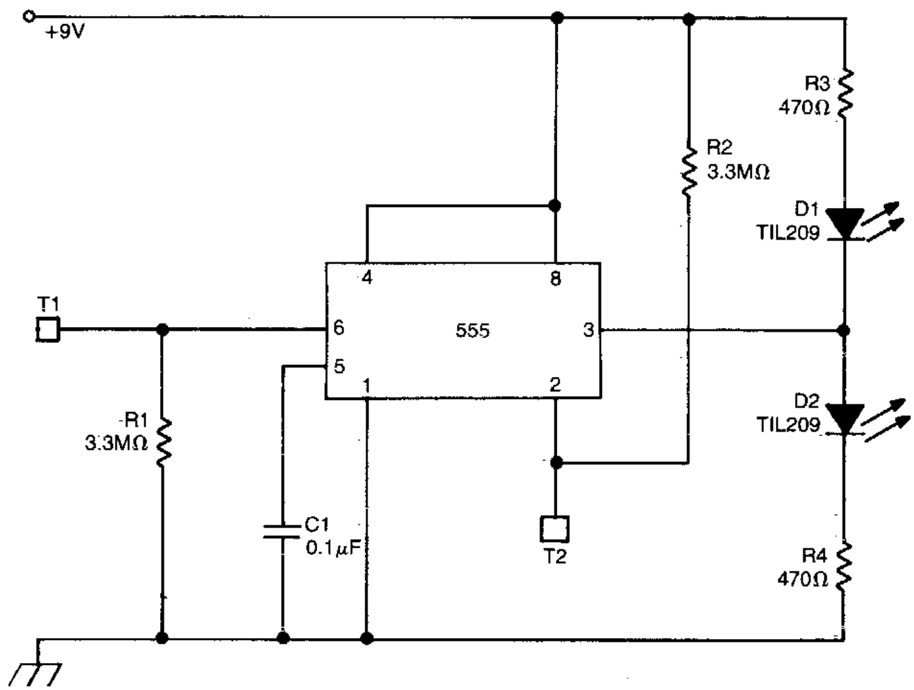


Fig. 14-7

Circuit Notes

This circuit uses a 555 timer in the bistable mode. Touching T2 causes the output to go high; D2 conducts and D1 extinguishes. Touching T1 causes the output to go low; D1 conducts and D2 is cut off. The output from pin 3 can also be used to operate other circuits (e.g., a triac controlled lamp). In this case, the LEDs are useful for finding the touch terminals in the dark. C1 is not absolutely necessary but helps to prevent triggering from spurious pulses.

CAPACITANCE OPERATED ALARM TO FOIL PURSE SNATCHERS

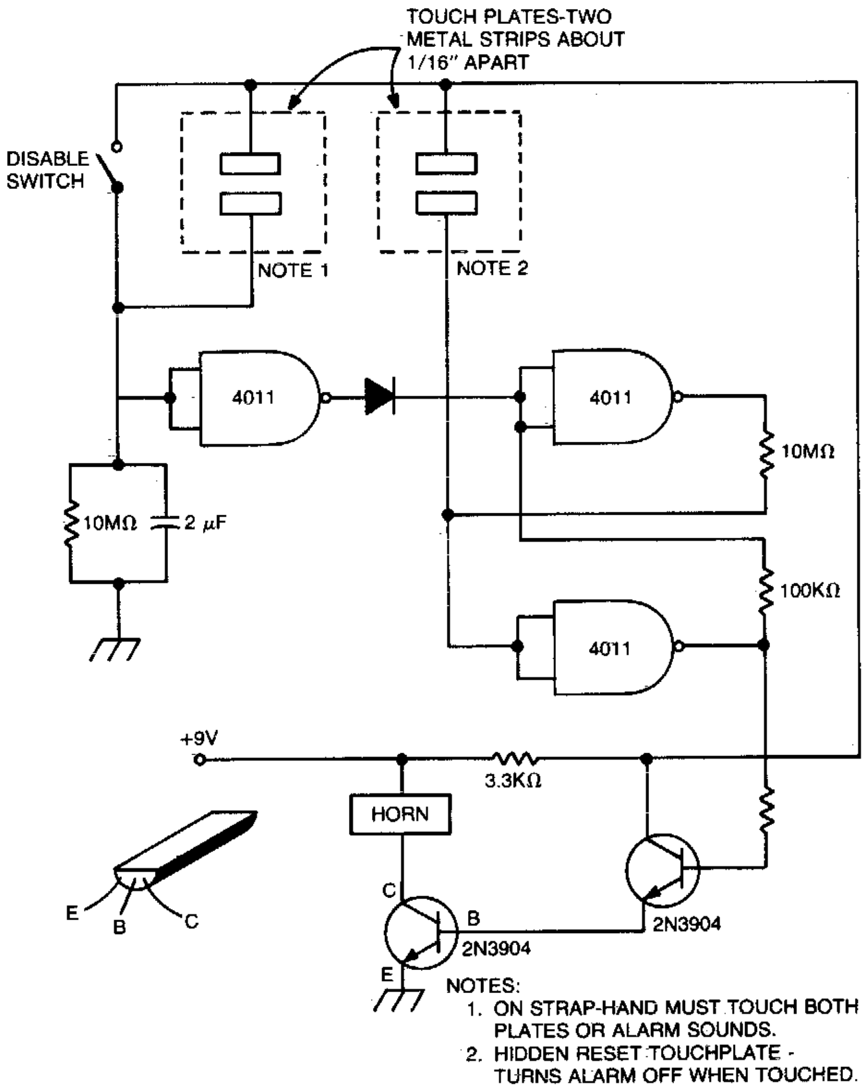
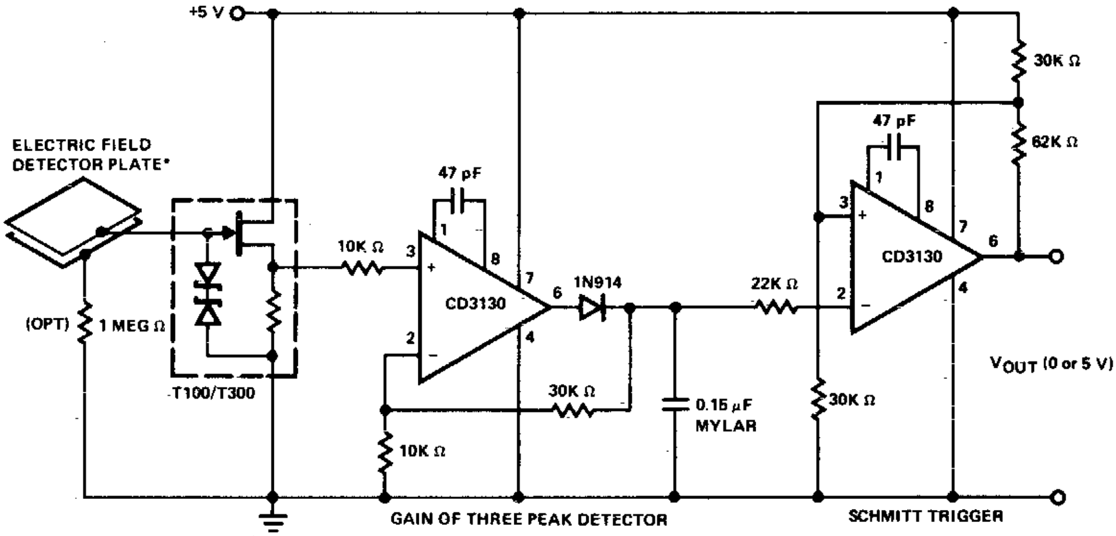


Fig. 14-8

Circuit Notes

As long as touch plates (1) are touched together, the alarm is off. If not held for about 30 seconds, the alarm goes off. The circuit can be disabled with switch or by touching the plates (2). The alarm is battery operated by a bicycle horn.

SELF-BIASED PROXIMITY SENSOR WORKS ON DETECTED CHANGING FIELD



*DETECTOR PLATE MAY BE DOUBLE-SIDED PC BOARD OR ANY INSULATED METAL SHEET

Fig. 14-9

TOUCH SWITCH OR PROXIMITY DETECTOR

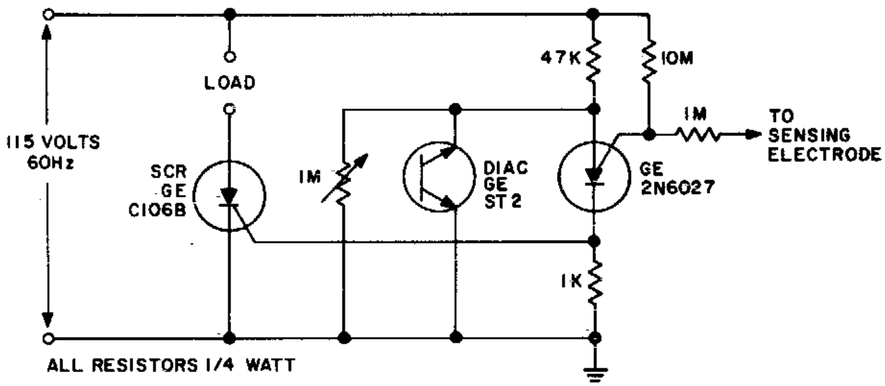


Fig. 14-10

Circuit Notes

This circuit is actuated by an increase in capacitance between a sensing electrode and the ground side of the line. The sensitivity can be adjusted to switch when a human body is within inches of the insulated plate used as the sensing electrode. Thus, sensitivity is adjusted with the 1 megohm potentiometer which determines the anode voltage level prior to clamping. This sensitivity will be proportional to the area of the surface opposing each other.

FINGER TOUCH OR CONTACT SWITCH

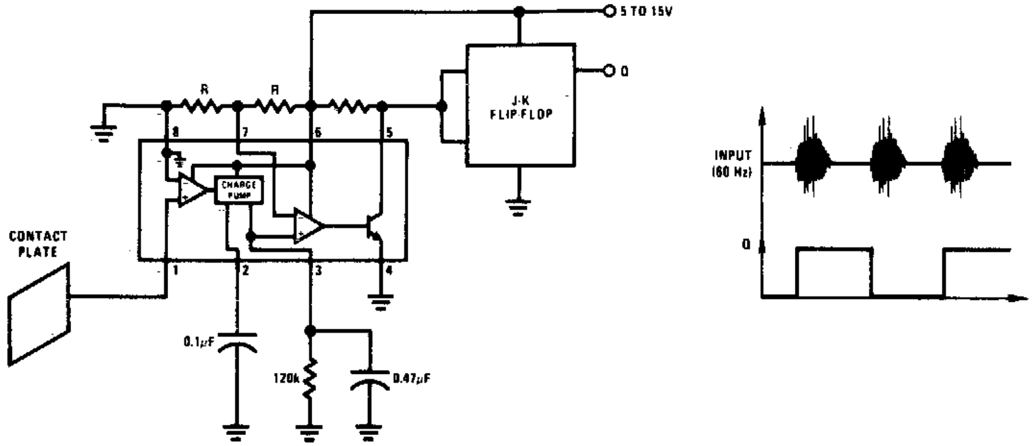


Fig. 14-11

PROXIMITY DETECTOR

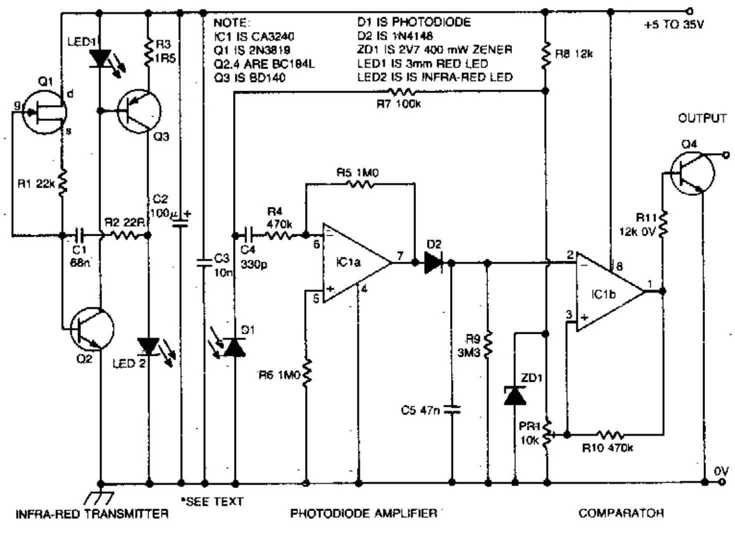


Fig. 14-12

Circuit Notes

The proximity sensor works on the principle of transmitting a beam of modulated infra-red light from the emitter diode LED2, and receiving reflections from objects passing in front of the beam with a photodiode detector D1. The circuit can be split into three distinct stages; the infra-red transmitter, the photodiode amplifier, and a variable threshold comparator.

TOUCH CIRCUIT

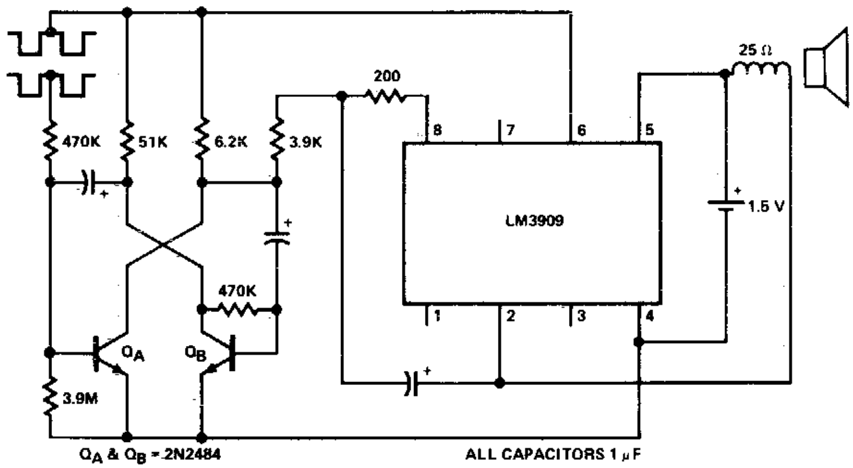


Fig. 14-13

CMOS TOUCH SWITCH

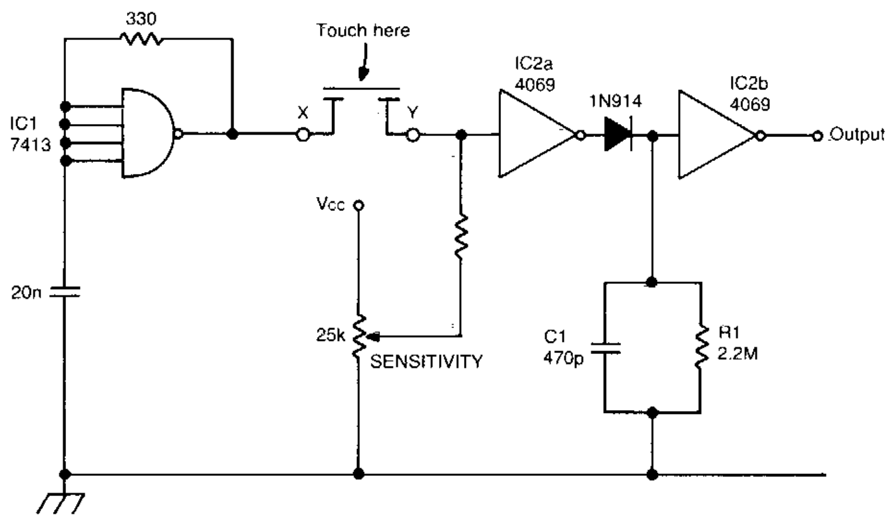


Fig. 14-14

Circuit Notes

This touch switch does not rely on mains hum for switching. It can be used with battery powered circuits. Schmitt trigger IC1 forms a 100 kHz oscillator and IC2a which is biased into the linear region, amplifies the output and charges C1 via the diode. IC2b acts as a level detector. When the sensor is touched, the oscillator signal is severely attenuated which causes C1 to discharge and IC2b to change state.

LATCHING, DOUBLE BUTTON TOUCH SWITCH

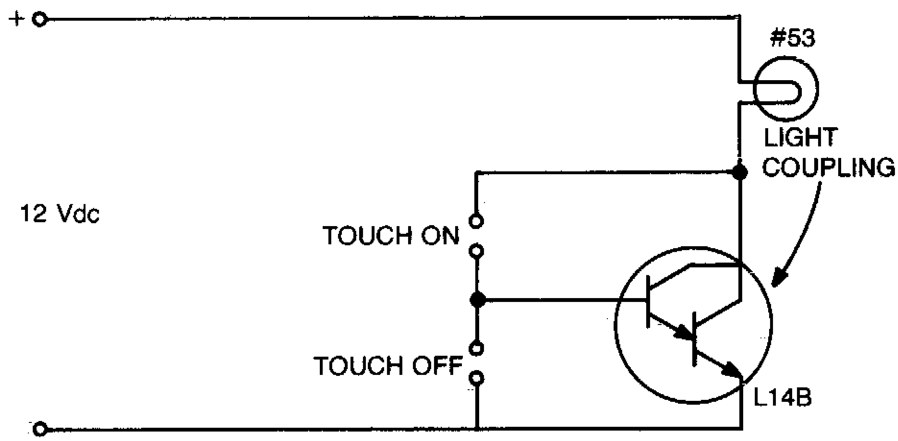


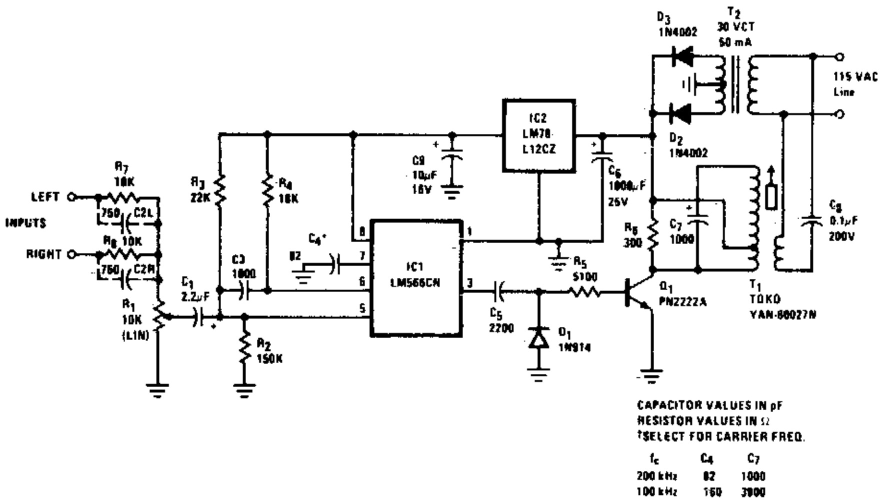
Fig. 14-15

15

Carrier Current Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Carrier System Transmitter



Carrier System Receiver

FM CARRIER CURRENT REMOTE SPEAKER SYSTEM

Circuit Notes

High quality, noise free, wireless FM transmitter/receiver operates over standard power lines. Complete system is suitable for high-quality transmission of speech or music, and will operate from any ac outlet anywhere on a one-acre homesite. Frequency response is 20-20, 000 Hz and THD is under ½%. Transmission distance along a power line is at least adequate to include all outlets in and around a suburban home and yard.

Two input terminals are provided so that both left and right signals of a stereo set may be combined for mono transmission to a single remote speaker if desired.

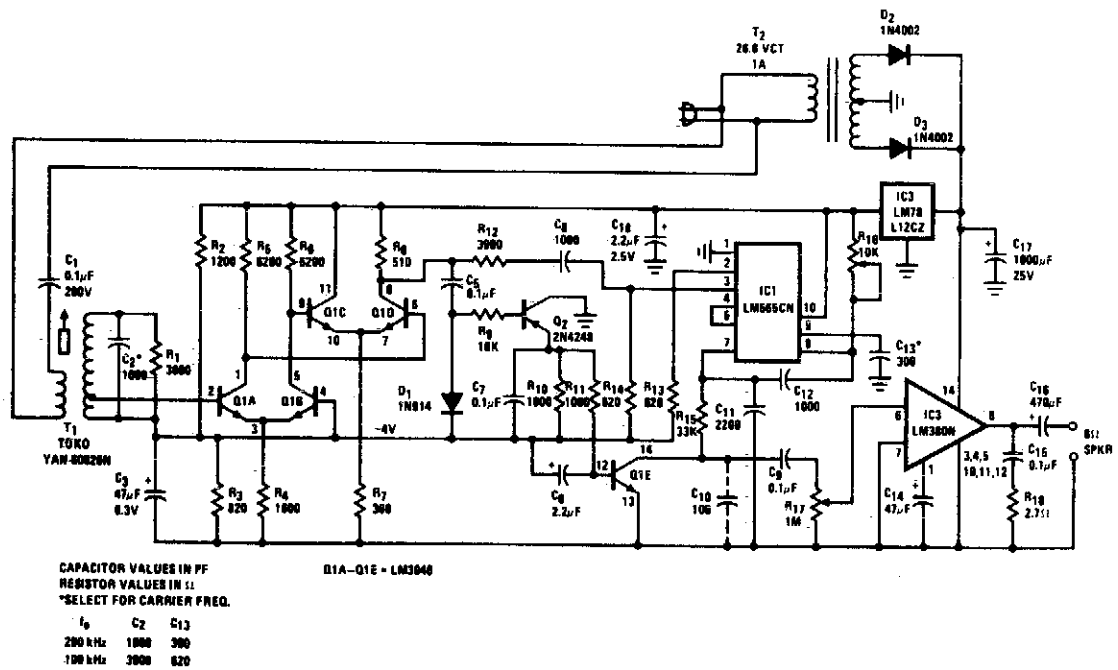
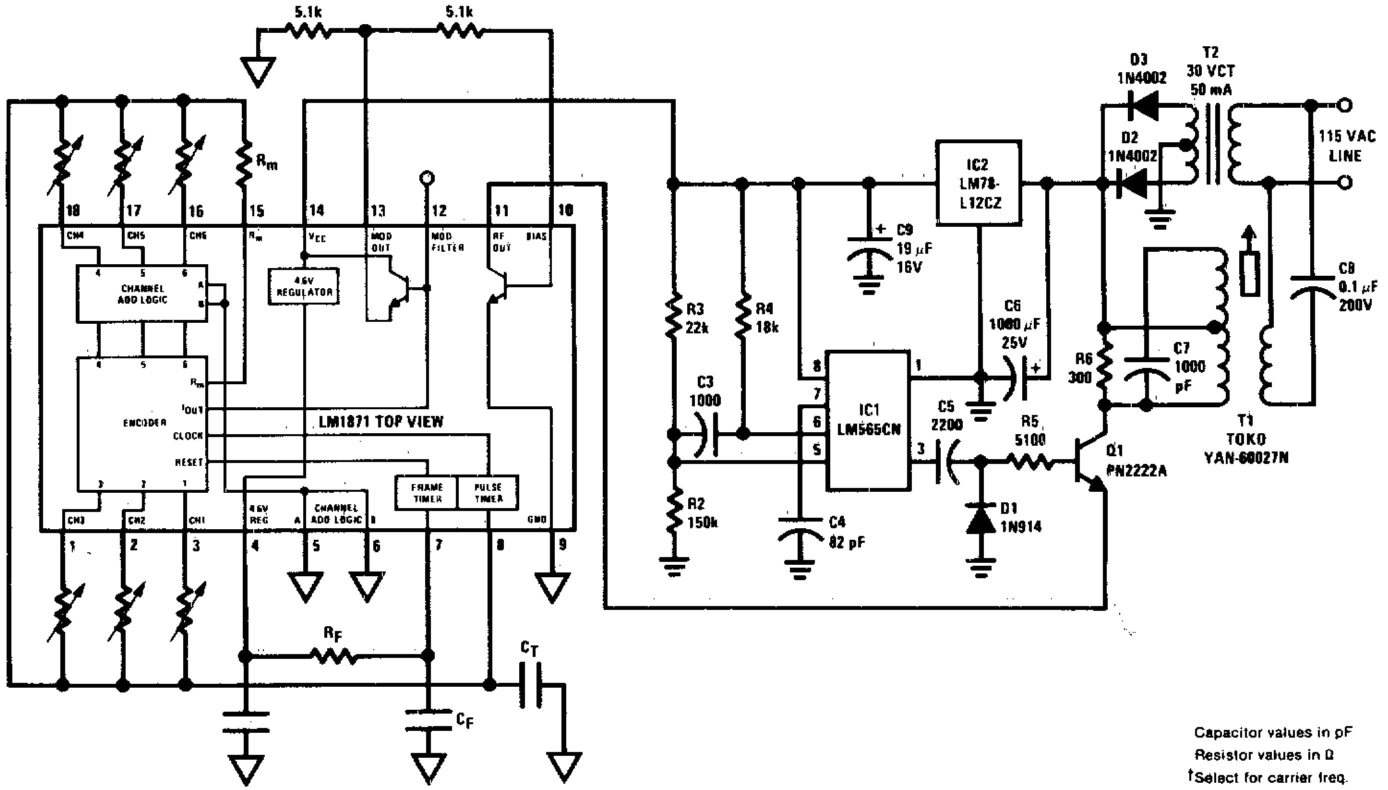


Fig. 15-1

The receiver amplifies, limits, and demodulates the received FM signal. It provides audio mute in the absence of carrier and 2.5 W output to a speaker.

200 kHz LINE CARRIER
TRANSMITTER WITH ON/OFF CARRIER MODULATOR



Capacitor values in pF
Resistor values in Ω
†Select for carrier freq.

f_c	C4	C7
200 kHz	82	1000
100 kHz	160	3900

Fig. 15-2

CARRIER CURRENT RECEIVER

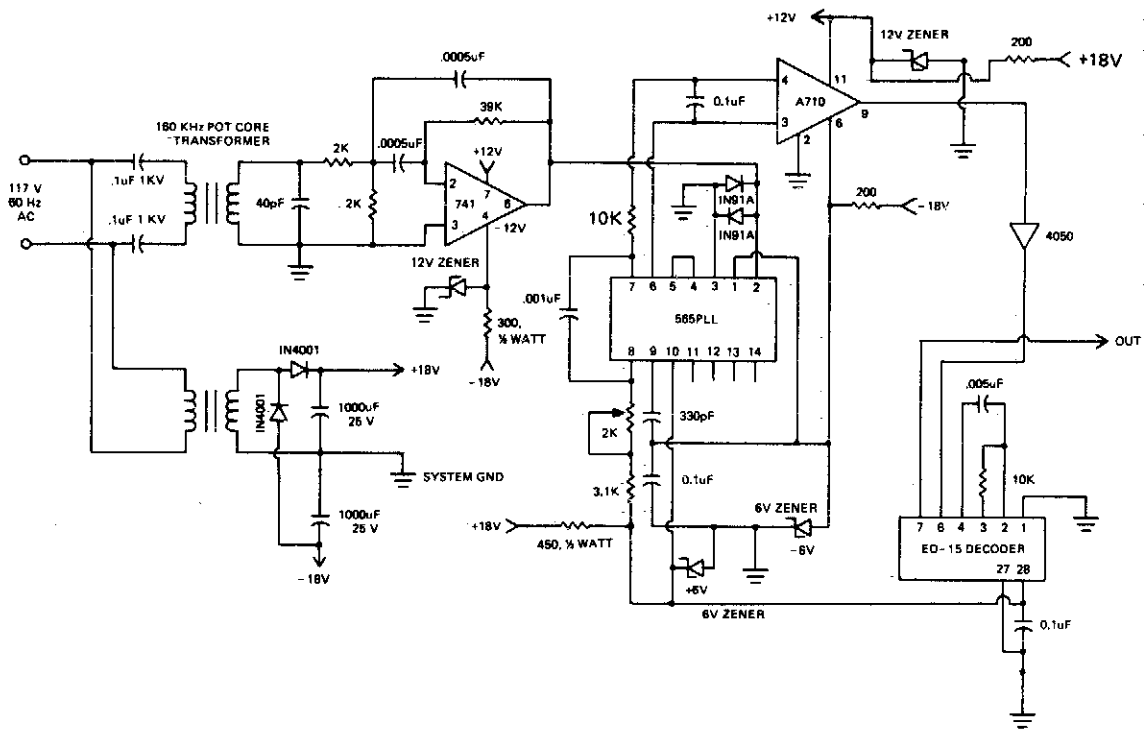


Fig. 15-3

Circuit Notes

160 kHz transformer consists of a 18 × 11mm ungapped pot core (Siemens, Ferrocube, etc.), utilizing magnetics incorporated type "F" material wound with 80½ turns of No. 35 wire for the secondary and 5½ turns for the primary. This gives a turns ratio of approximately 15 to 1.

CARRIER CURRENT TRANSMITTER

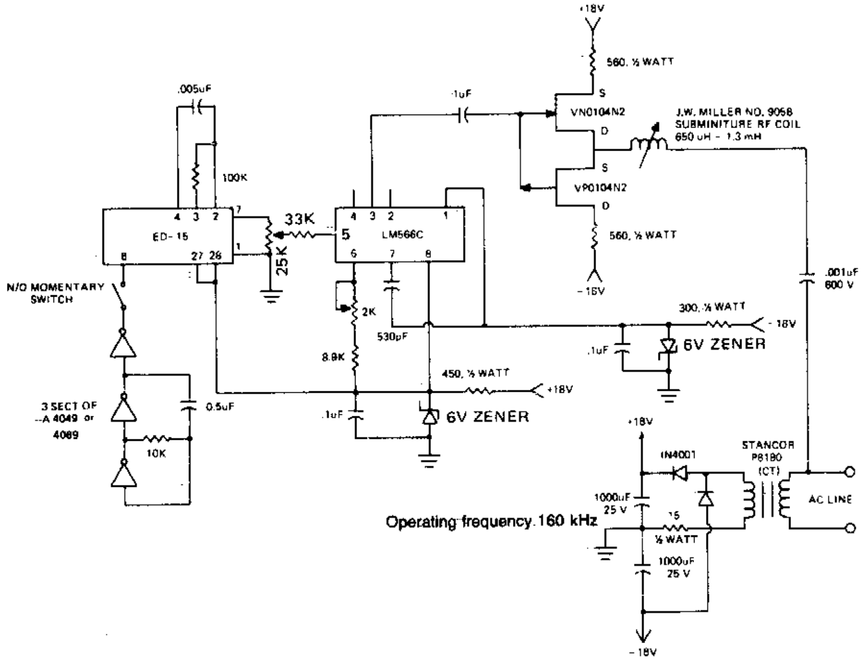


Fig. 15-4

CARRIER CURRENT TRANSMITTER

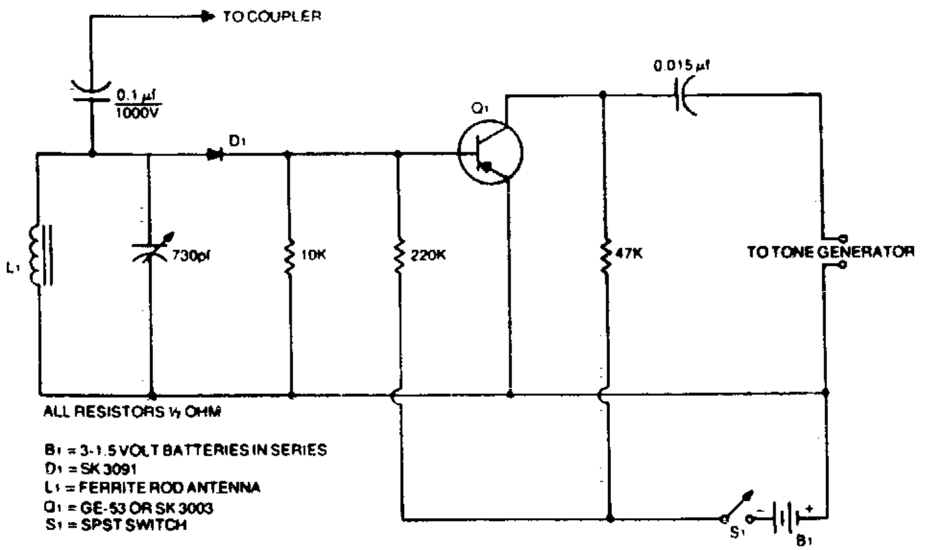


Fig. 15-5

IC CARRIER CURRENT TRANSMITTER

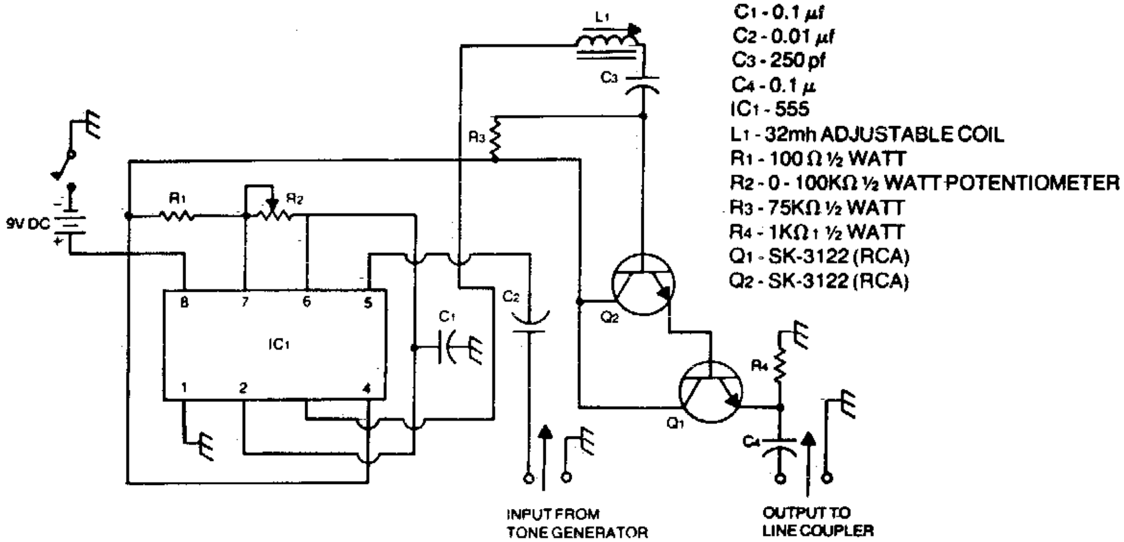


Fig. 15-6

SINGLE TRANSISTOR CARRIER CURRENT RECEIVER

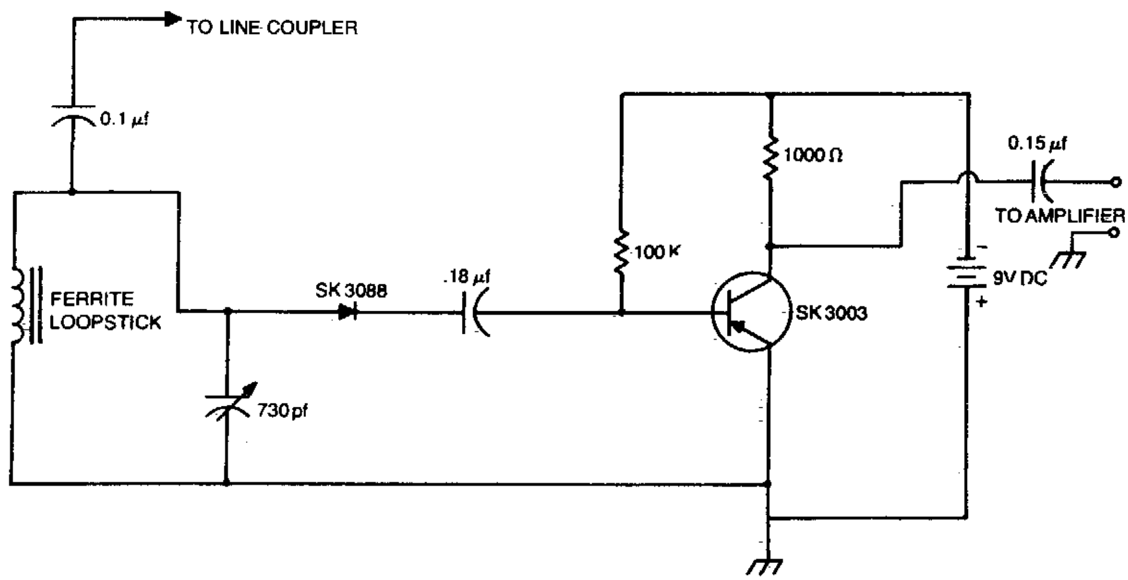
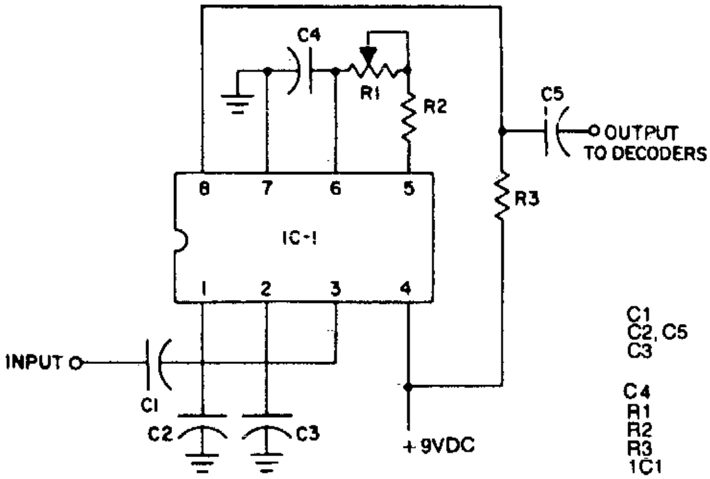


Fig. 15-7

IC CARRIER-CURRENT RECEIVER



C1	0.1 μF, 600V PAPER CAPACITOR
C2, C5	0.1 μF, 50V MYLAR CAPACITOR
C3	22μF, 35 VOLT ELECTROLYTIC CAPACITOR
C4	.0047 μF25 VOLT CAPACITOR
R1	20K POTENTIOMETER
R2	2K, ½ WATT RESISTOR
R3	4K, ½ WATT RESISTOR
1C1	TYPE 567 TONE DECODER

Fig. 15-8

CARRIER-CURRENT REMOTE CONTROL OR INTERCOM

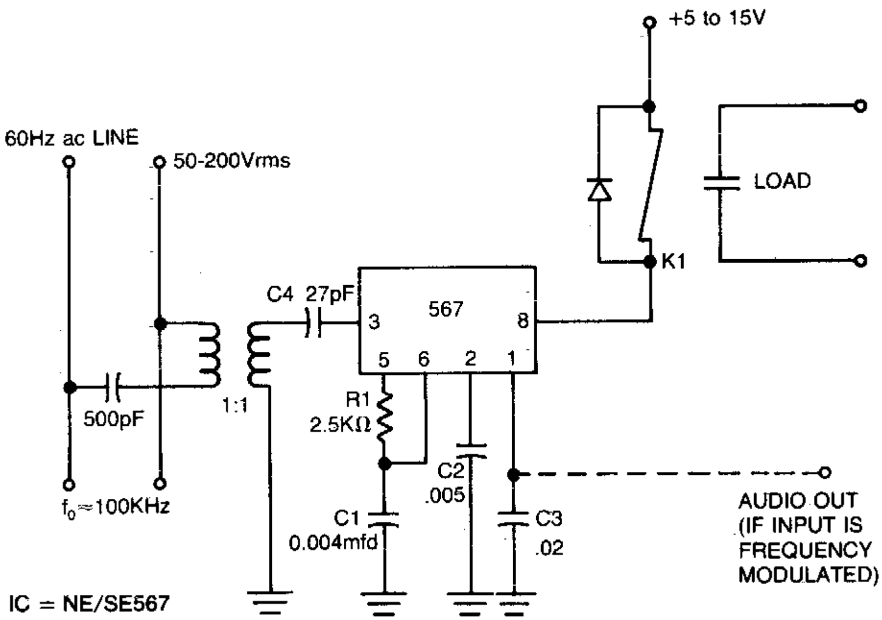


Fig. 15-9

16

Comparators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

NULL DETECTOR

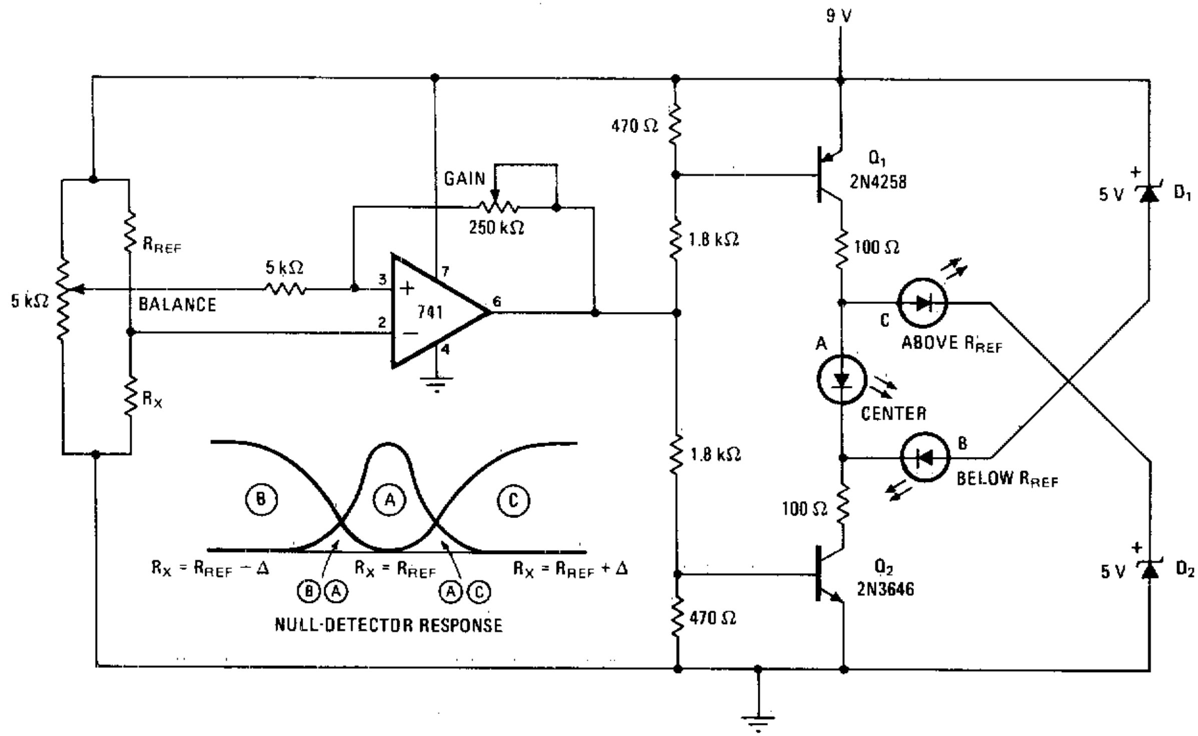


Fig. 16-1

Circuit Notes

Null detector uses simple LED readout to indicate if test resistor R_x is below, equal to, or greater than test resistance R_{ref}. If $R_x = R_{ref}$, the 741 output sits at midpoint value of 4.5 volts and LED A lights. Otherwise, the output of the 741 turns off one transistor and diverts current from the other transistor through B or C, depending on the polarity of the input voltage difference. Null-detector response is illustrated.

COMPARATOR WITH VARIABLE HYSTERESIS (WITHOUT SHIFTING INITIAL TRIP POINT)

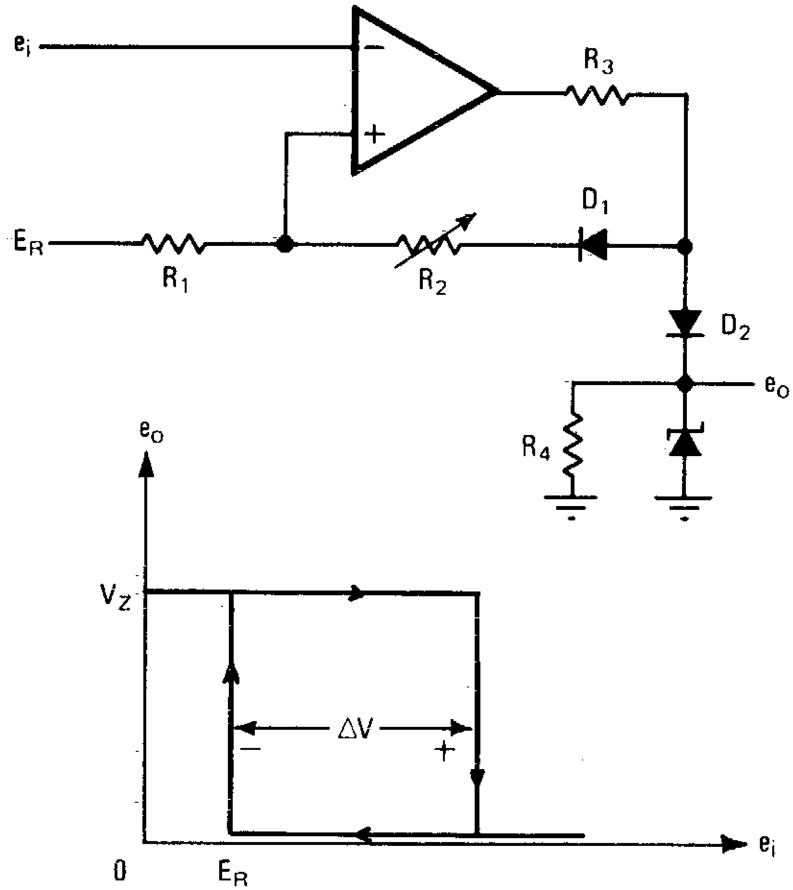
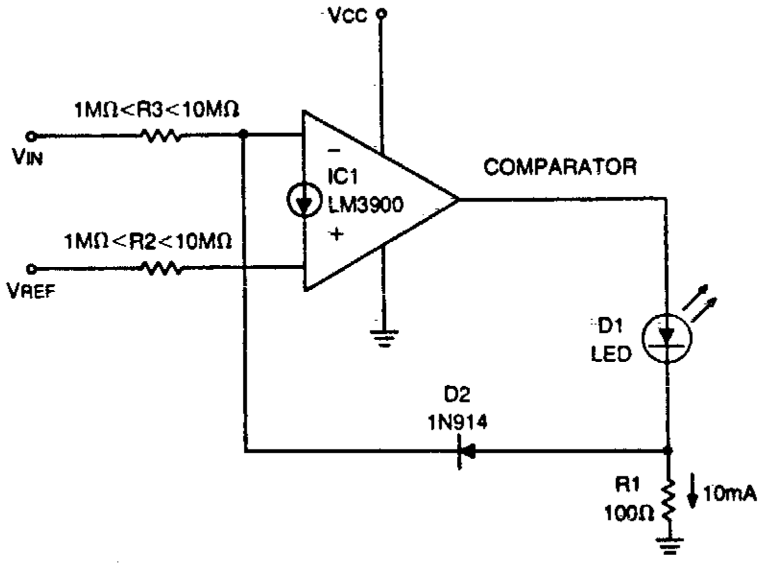


Fig. 16-2

Circuit Notes

An operational amplifier can be used as a convenient device for analog comparator applications that require two different trip points. The addition of a positive-feedback network introduces a precise variable hysteresis into the usual comparator switching action. Such feedback develops two comparator trip points centered about the initial trip point or reference point. The voltage difference, ΔV, between the trip points can be adjusted by varying resistor R2. When the output voltage is taken from the zener diode, as shown, it switches between zero and V_Z, the zener voltage.

DIODE FEEDBACK COMPARATOR



Parts list
IC1—LM3900
D1—LED Lafayette 32P06331V
D2—IN914
All resistors 1/4W
R1—100Ω
R2—(See circuit) 1MΩ to 10MΩ
R3—(See circuit) 1MΩ to 10MΩ

Fig. 16-3

Circuit Notes

This circuit can drive an LED display with constant current independently of wide power supply voltage changes. It can operate with a power supply range of at least 4V to 30V. With 10M resistances for R2 and R3 and the inverting input of the comparator grounded, the circuit becomes an LED driver with very high input impedance. The circuit can also be used in many other applications where a controllable constant current source is needed.

UNDERVOLTAGE/OVERVOLTAGE INDICATOR

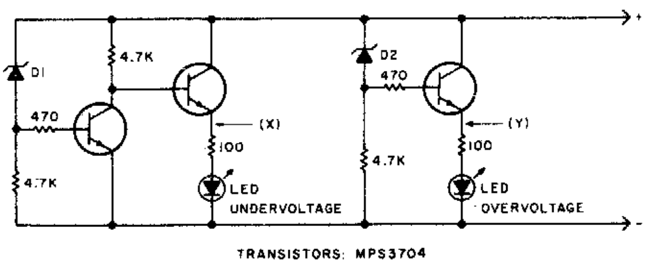


Fig. 16-4

Circuit Notes

This circuit will make the appropriate LED glow if the monitored voltage goes below or above the value determined by zener diodes D1 and D2.

DUAL LIMIT COMPARATOR

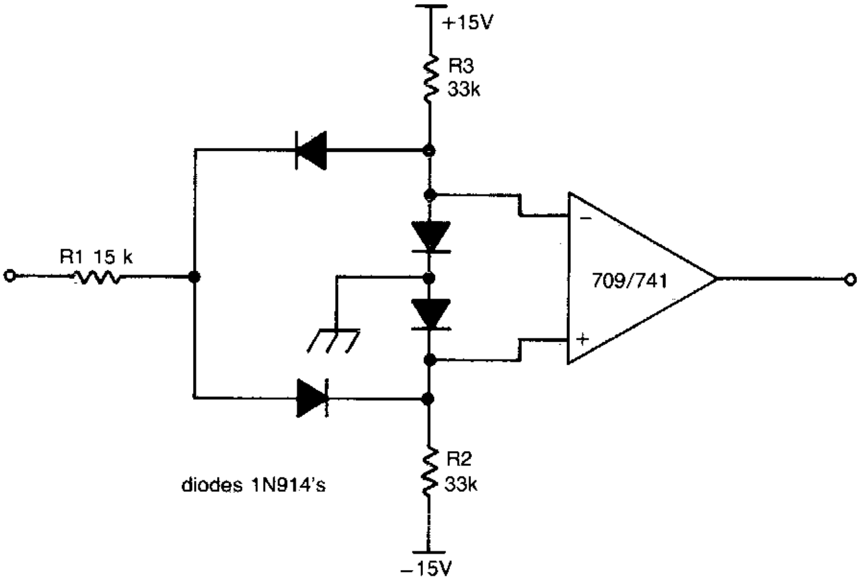


Fig. 16-5

Circuit Notes

This circuit gives a positive output when the input voltage exceeds 8.5 volts. Between these limits the output is negative. The positive limit point is determined by the ratio of R1, R2, and the negative point by R1, R3. The forward voltage drop across the diodes must be allowed for. The output may be inverted by reversing the inputs to the op amp. The 709 is used without frequency compensation.

HIGH/LOW LIMIT ALARM

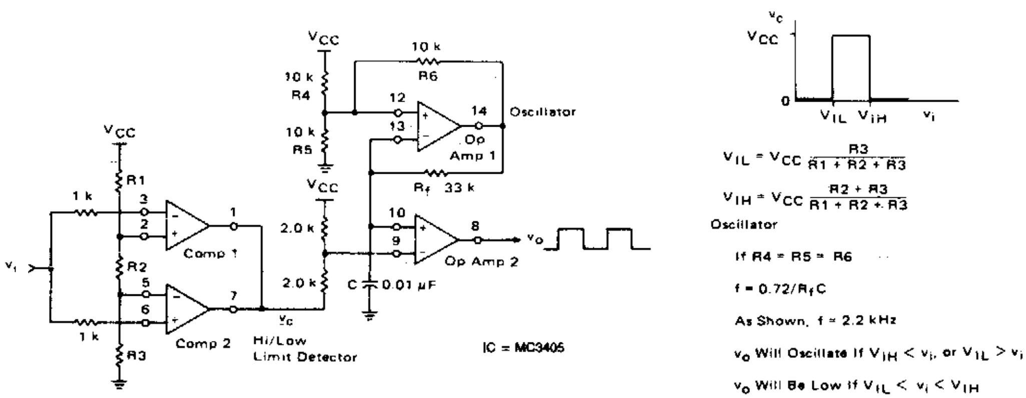
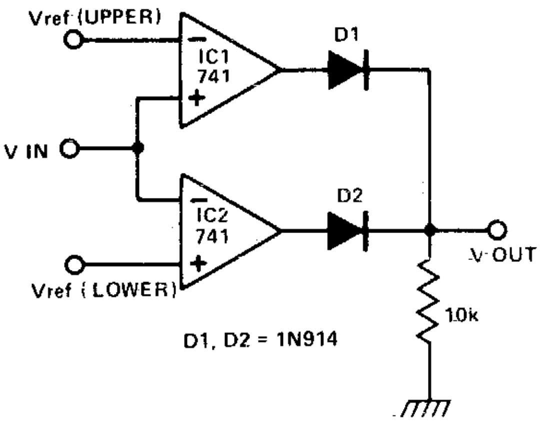


Fig. 16-6

WINDOW COMPARATOR



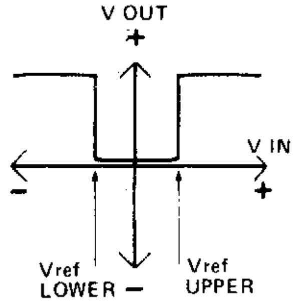
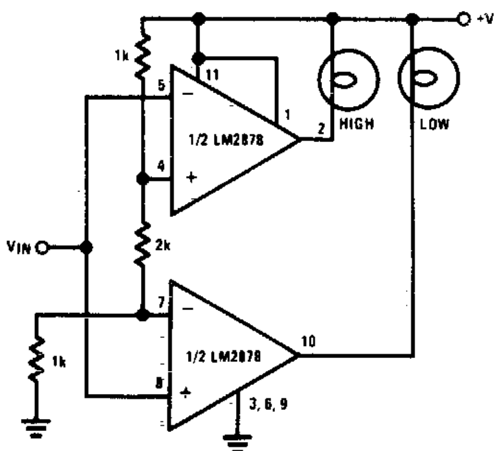


Fig. 16-7

Circuit Notes

This circuit gives an output (which in this case is 0V) when an input voltage lies in between two specified voltages. When it is outside this window, the output is positive. The two op amps are used as voltage comparators. When Vin is more positive than Vref (upper) the output of IC1 is positive and D1 is forward biased. Otherwise the output is negative, D1 reverse biased and hence Vout is 0V. Similarly, when Vin is more negative than Vref (lower), the output of IC2 is positive; D2 is forward biased and this Vout is positive. Otherwise Vout is 0V. When Vin lies within the window set by the reference voltages, Vout is 0V.

WINDOW COMPARATOR DRIVING HIGH/LOW LAMPS



TRUTH TABLE

V_{IN}	High	Low
< 1/4 V+	Off	On
1/4 V+ to 3/4 V+	Off	Off
> 3/4 V+	On	Off

Fig. 16-8

COMPARATOR WITH TIME OUT

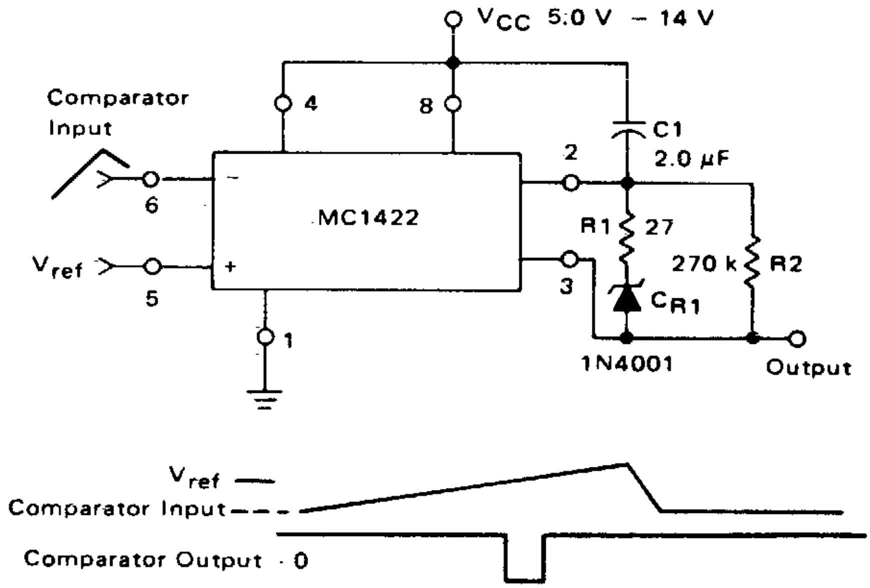


Fig. 16-9

Circuit Notes

The MC1422 is used as a comparator with the capability of a timing output pulse when the inverting input (Pin 6) is ≥ the noninverting input (Pin 5). The frequency of the pulses for the values of R2 and C1 as shown is approximately 2.0 Hz, and the pulse width 0.3 ms.

NONINVERTING COMPARATOR WITH HYSTERESIS

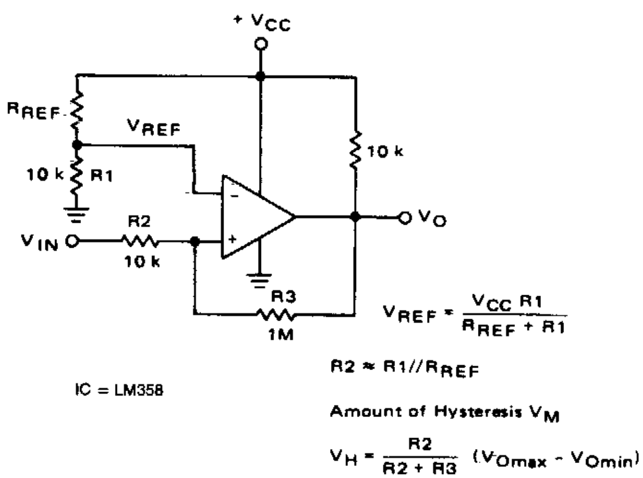
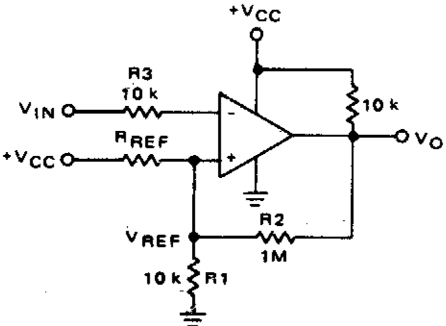


Fig. 16-10

INVERTING COMPARATOR WITH HYSTERESIS



$$V_{REF} \approx \frac{V_{CC}\,R1}{R_{REF} + R1}$$

$$R3 \approx R1 \,//\, R_{REF} \,//\, R1$$

$$V_H = \frac{R1 // R_{REF}}{R1 // R_{REF} + R2} \quad (V_{Omax} - V_{Omin})$$

Fig. 16-11

WINDOW COMPARATOR

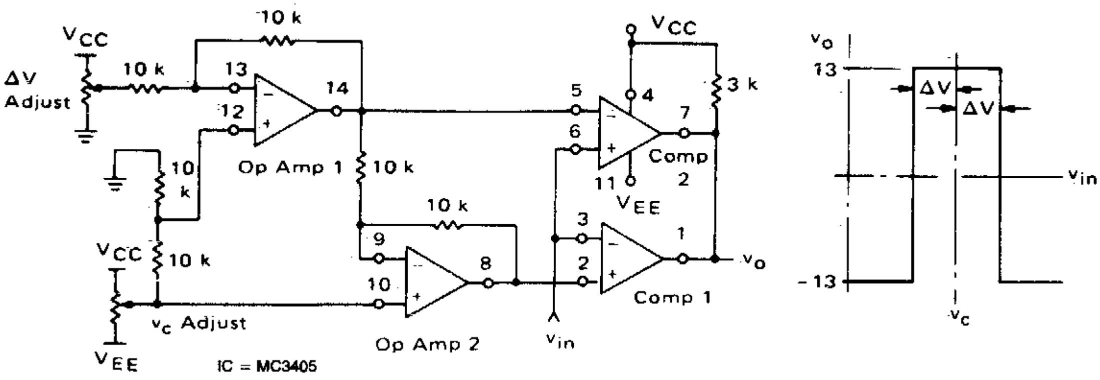


Fig. 16-12

MICROPOWER DOUBLE-ENDED LIMIT DETECTOR

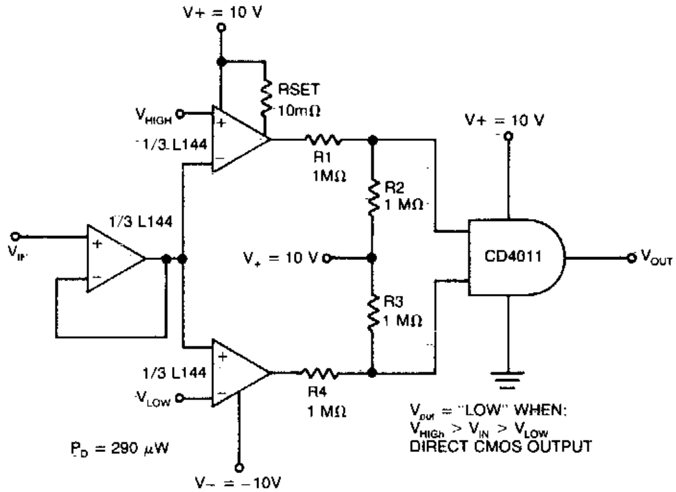


Fig. 16-13

Circuit Notes

The detector uses three sections of an L144 and a DC4011 type CMOS NAND gate to make a very low power voltage monitor. If the input voltage, V_{IN}, is above V_{HIGH} or below V_{LOW}, the output will be a logical high. If (and only if) the input is between the limits will the output be low. The 1 megohm resistors R1, R2, R3, and R4 translate the bipolar ±10V swing of the op amps to a 0 to 10V swing acceptable to the ground-referenced CMOS logic.

OPPOSITE POLARITY INPUT VOLTAGE COMPARATOR

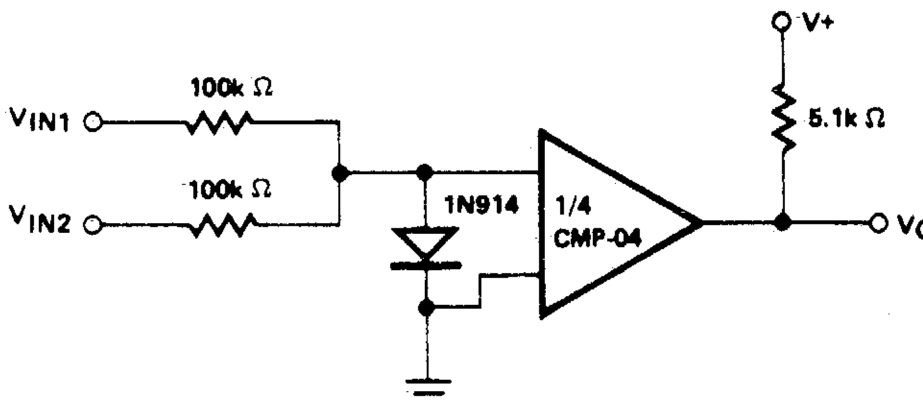


Fig. 16-14

LIMIT COMPARATOR

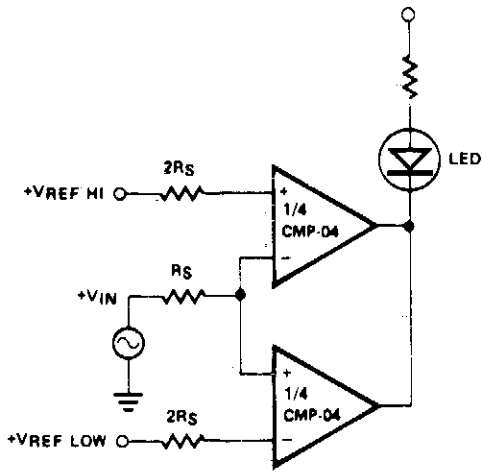


Fig. 16-15

DOUBLE-ENDED LIMIT COMPARATOR

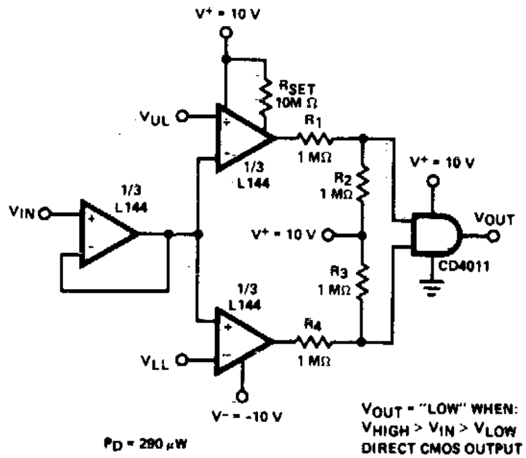


Fig. 16-17

COMPARATOR CLOCK CIRCUIT

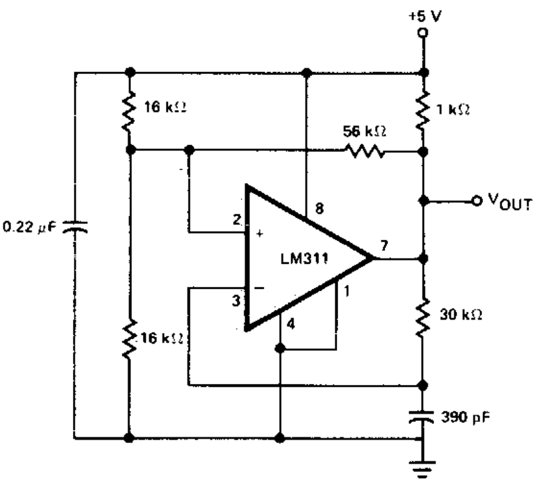


Fig. 16-16

LIMIT COMPARATOR

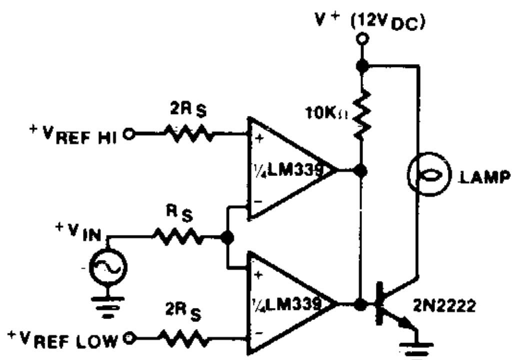


Fig. 16-18

PRECISION, DUAL LIMIT, GO/NO GO TESTER

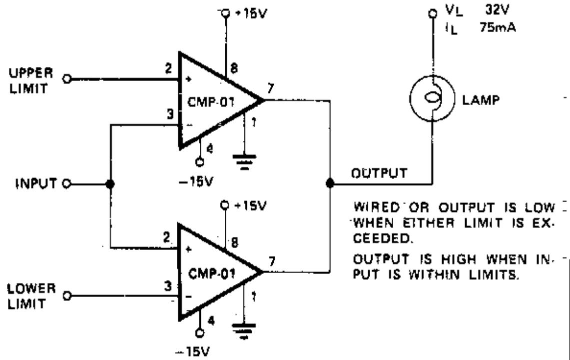
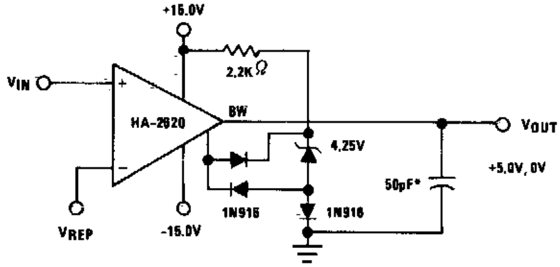


Fig. 16-19

HIGH IMPEDANCE COMPARATOR

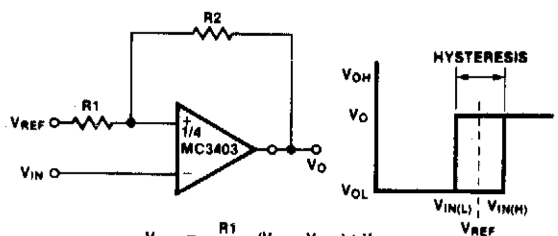


IC = HA-2620

Harris Semiconductor
Analog Data Book - 1984

Fig. 16-21

COMPARATOR WITH HYSTERESIS



$$V_{IN(L)} = \frac{R1}{R1 + R2}(V_{OL} - V_{REF}) + V_{REF}$$

$$V_{IN(H)} = \frac{R1}{R1 + R2}(V_{OH} - V_{REF}) + V_{REF}$$

$$H = \frac{R1}{R1 + R2}(V_{OH} - V_{OL})$$

Fig. 16-20

COMPARATOR

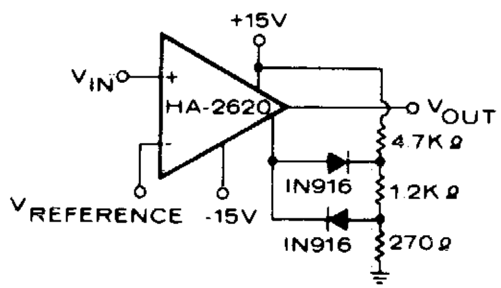


Fig. 16-22

Circuit Notes

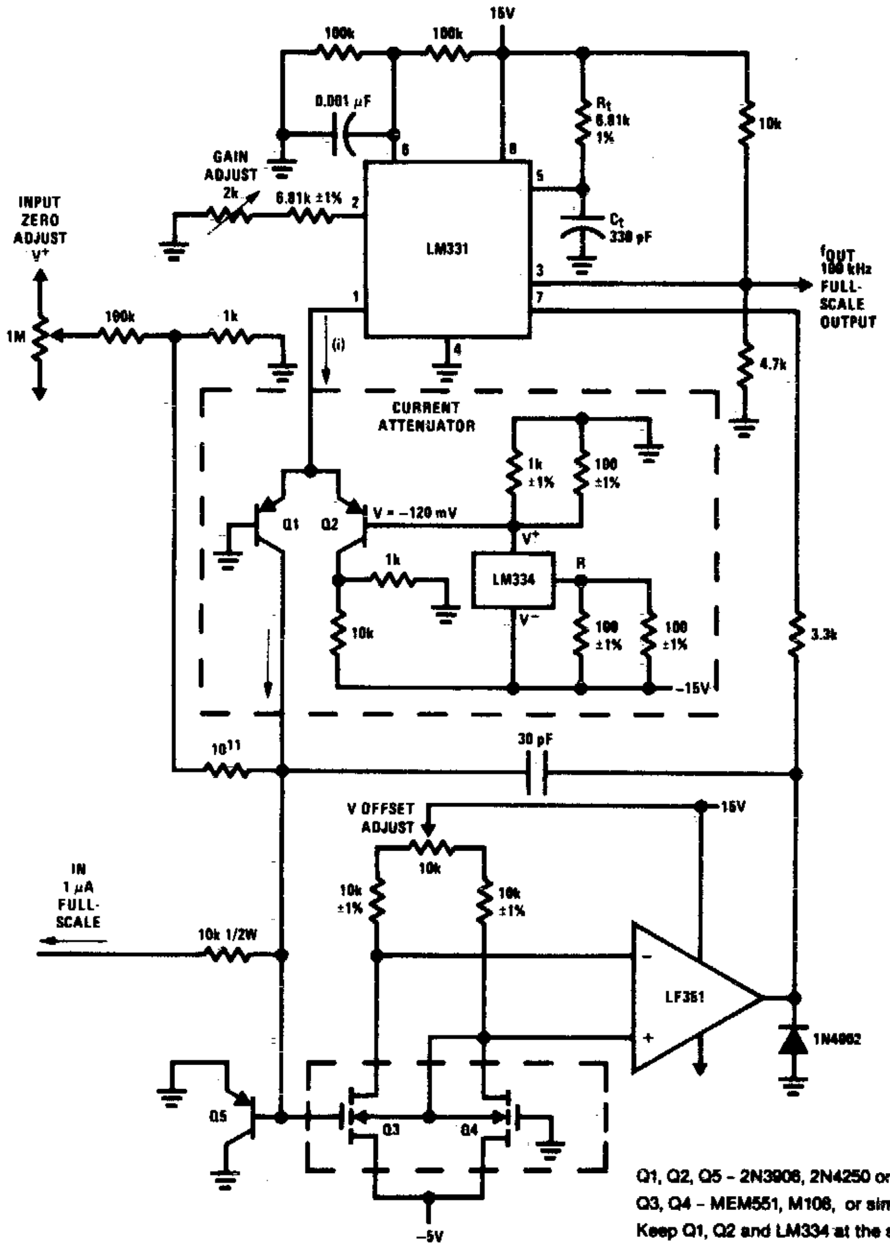
An operational amplifier is used as a comparator which is capable of driving approximately 10 logic gates.

17

Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PICOAMPERE-TO-FREQUENCY CONVERTERS



Q1, Q2, Q5 - 2N3906, 2N4250 or similar
Q3, Q4 - MEM551, M106, or similar. See text
Keep Q1, Q2 and LM334 at the same temperature

Fig. 17-1

BCD-TO-ANALOG CONVERTER

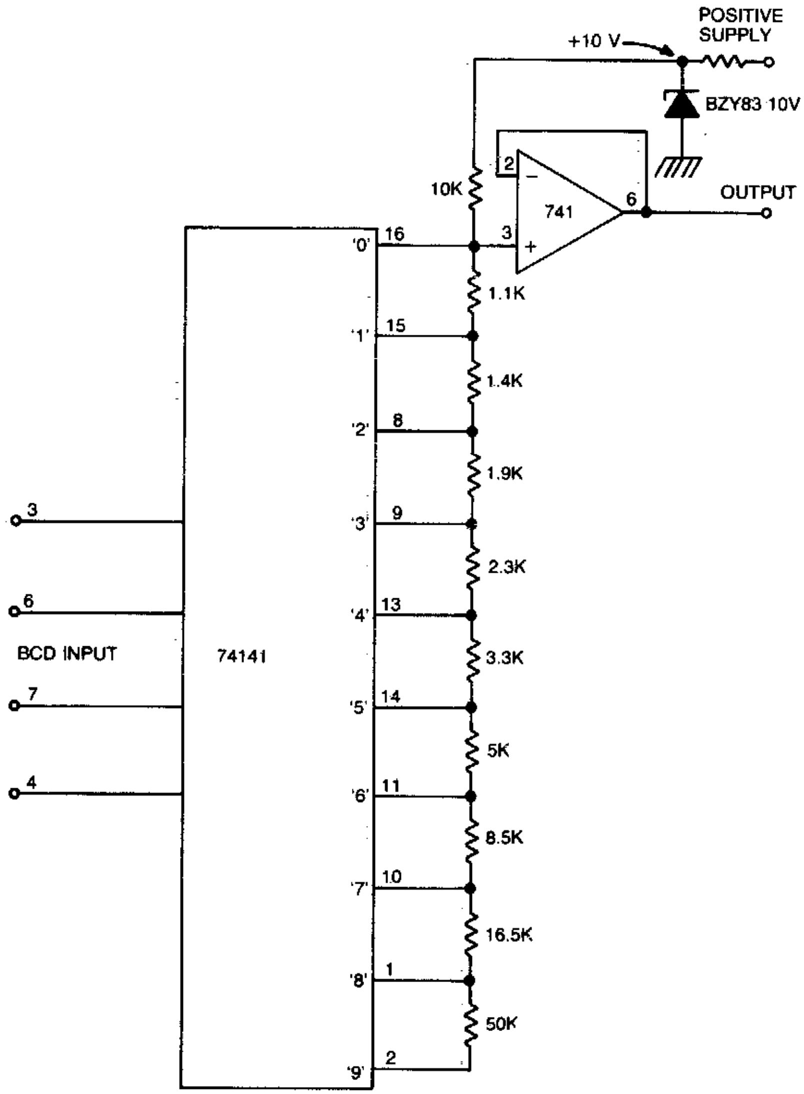


Fig. 17-2

Circuit Notes

This circuit will convert four-bit BCD into a variable voltage from 0-9 V in 1 V steps. The SN74141 is a Nixie driver, and has ten open-collector outputs. These are used to ground a selected point in the divider chain determined by the BCD code at the input, and so produce a corresponding voltage at the output. Accuracy of the circuit depends on the tolerance of the resistors and the accuracy of the reference voltage. However, presets can be used in the divider chain, with correct calibration. The 741 is used as a buffer.

RESISTANCE-TO-VOLTAGE CONVERTER

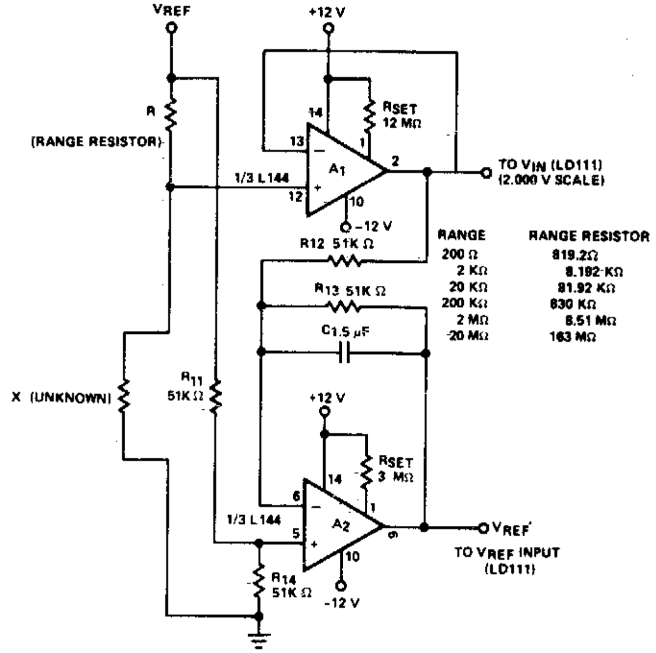


Fig. 17-3

Circuit Notes

Circuit will measure accurately to 20 M when associated with a buffer amplifier (A1) having a low input bias current (I_{IN}) < 30 nA). The circuit uses two of the three amplifiers contained in the Siliconix L144 micropower triple op amp.

LOW-COST, μP INTERFACED, TEMPERATURE-TO-DIGITAL CONVERTER

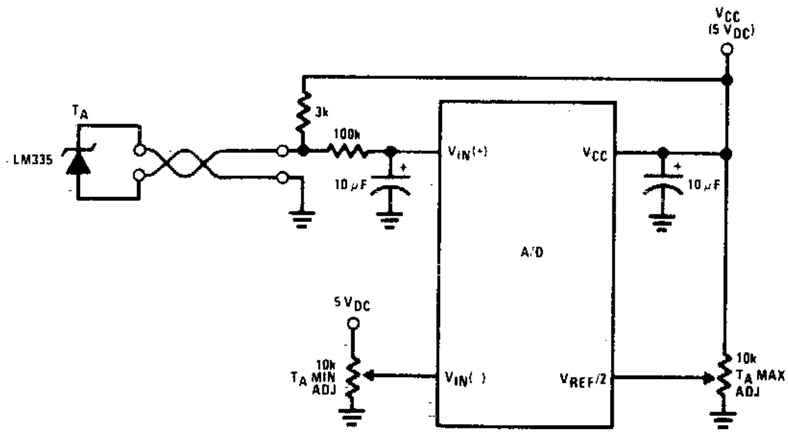


Fig. 17-4

HI-LO RESISTANCE-TO-VOLTAGE CONVERTER

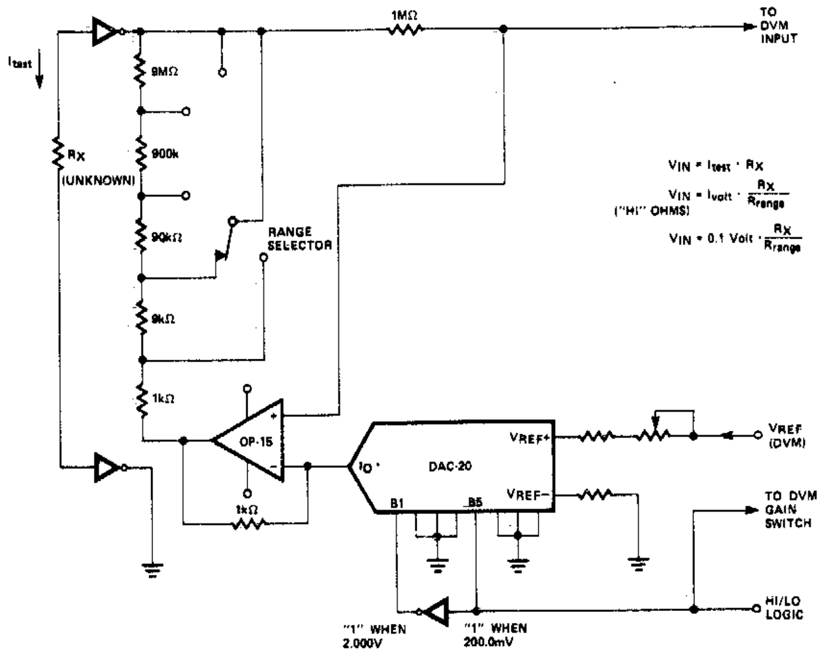
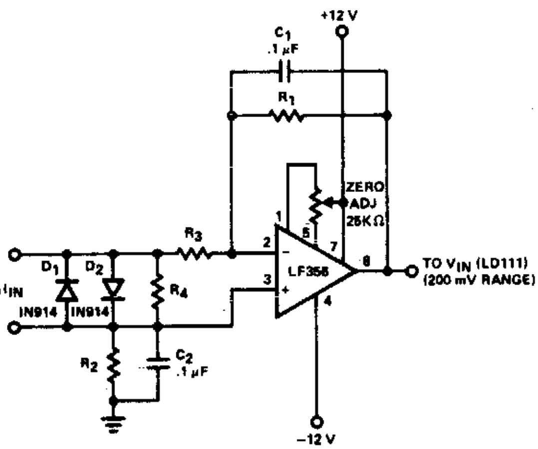


Fig. 17-5

CURRENT-TO-VOLTAGE CONVERTER



CURRENT RANGE	R_1	R_2	R_3	R_4
200 nA	500 KΩ	500 KΩ	0	∞
2 μA	50 KΩ	50 KΩ	0	∞
20 μA	5 KΩ	5 KΩ	0	∞
200 μA	1 KΩ	0	0	∞
2 mA	50 KΩ	0	5.0 K	10.0 Ω
20 mA	50 KΩ	0	5.0 K	1.0 Ω
200 mA	50 KΩ	0	5.0 K	.1 Ω
2 A	50 KΩ	0	5.0 K	.01Ω

Circuit Notes

Converter features eight decades of current range. The circuit is intended to be used with the 200.0 mV range of a DVM.

Fig. 17-6

CALCULATOR-TO-STOPWATCH CONVERTER

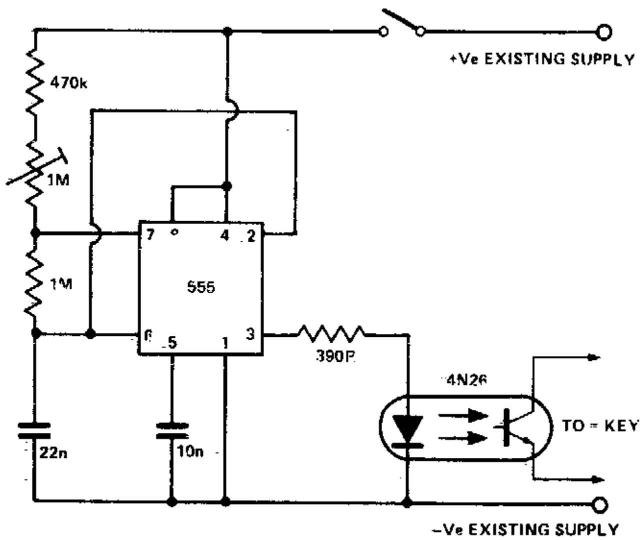
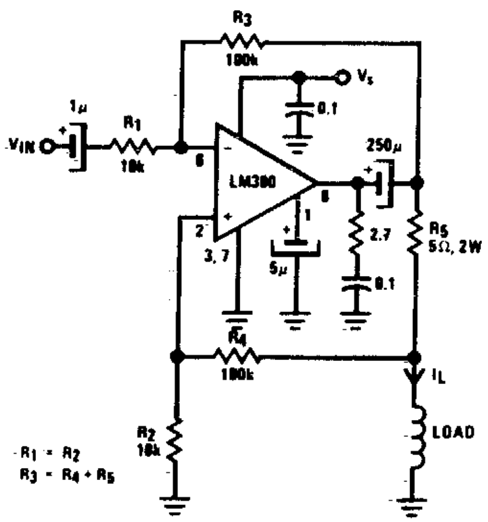


Fig. 17-7

Circuit Notes

This circuit can be fitted to any calculator with an automatic constant to enable it to be used as a stop-watch. The 555 timer is set to run at a suitable frequency and connected to the existing calculator battery via the push-on push-off switch and the existing calculator on-off switch.

POWER VOLTAGE-TO-CURRENT CONVERTER



Circuit Notes

Low cost converter is capable of supplying constant ac currents up to 1 A over variable loads.

Fig. 17-8

HIGH IMPEDANCE PRECISION RECTIFIER FOR AC/DC CONVERTER

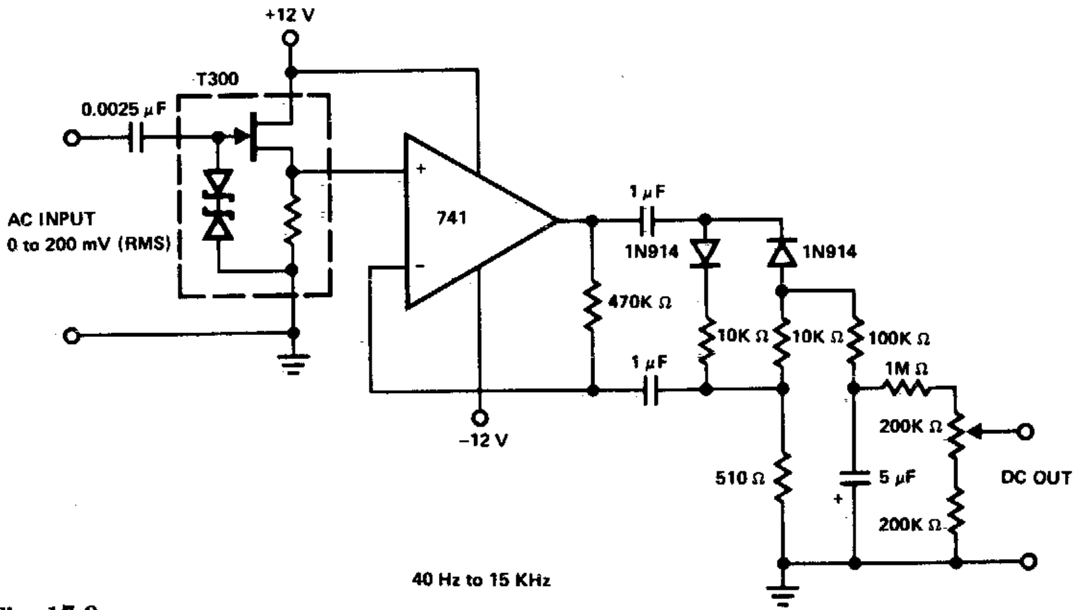


Fig. 17-9

WIDE-RANGE CURRENT-TO-FREQUENCY CONVERTER

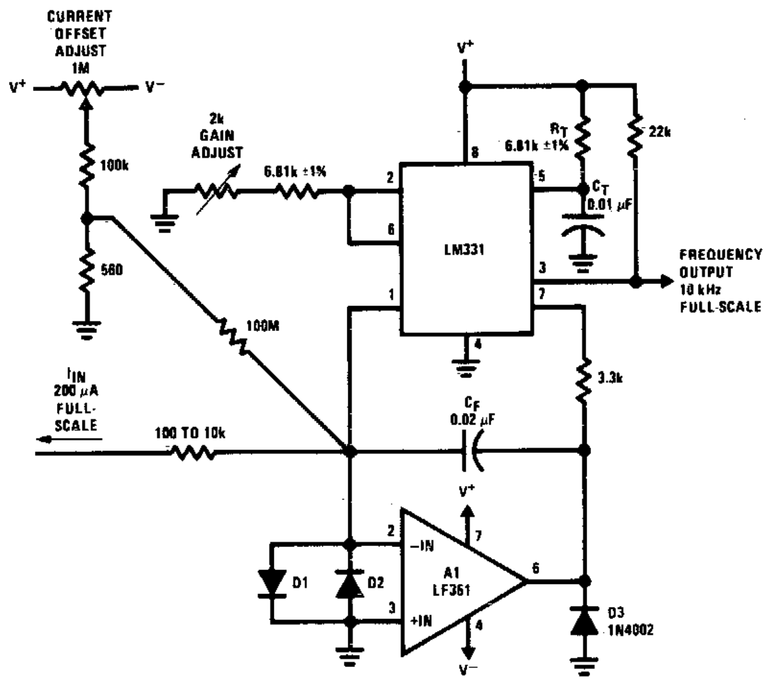


Fig. 17-10

AC-TO-DC CONVERTER

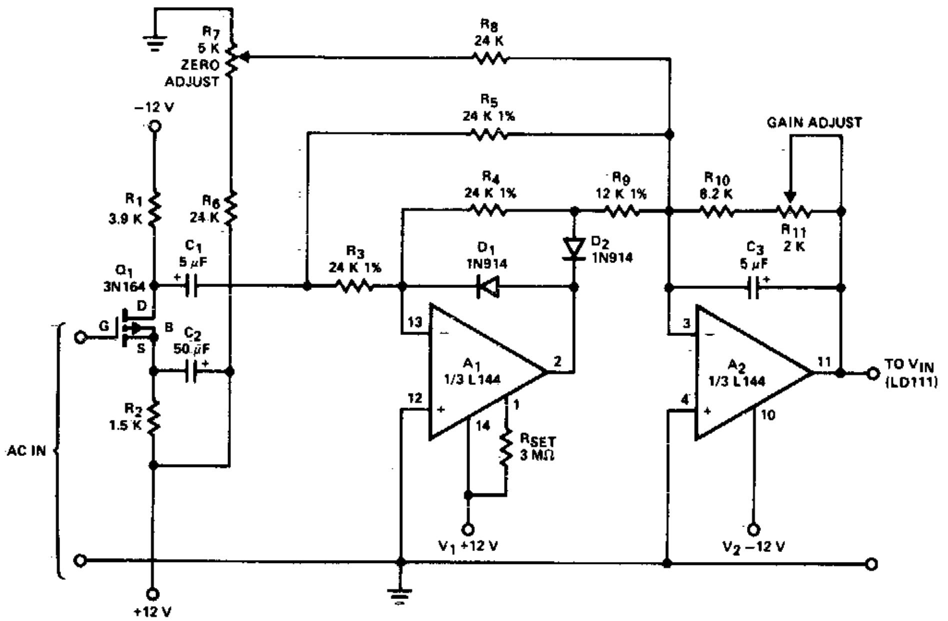


Fig. 17-11

Circuit Notes

This circuit includes a PMOS enhancement-mode FET input buffer amplifier, coupled to a classical absolute value circuit which essentially eliminates the effect of the forward voltage drop across diodes D1 and D2.

CURRENT-TO-VOLTAGE CONVERTER WITH 1% Accuracy

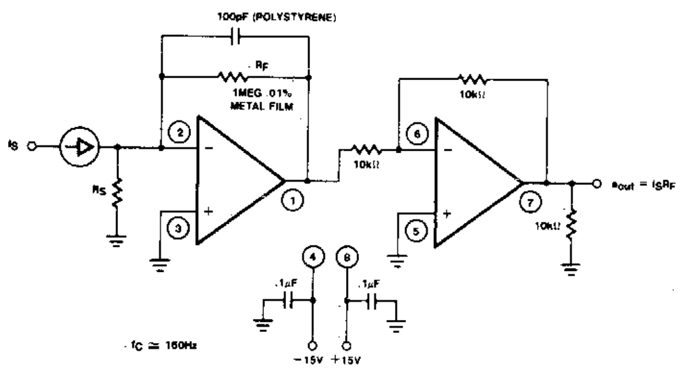


Fig. 17-12

IC = NE/SE55/1
(Sensitivity: 1 Volt per Microamp)

Circuit Notes

A filter removes the dc component of the rectified ac, which is then scaled to RMS. The output is linear from 40 Hz to 10 kHz or higher.

POLARITY CONVERTER

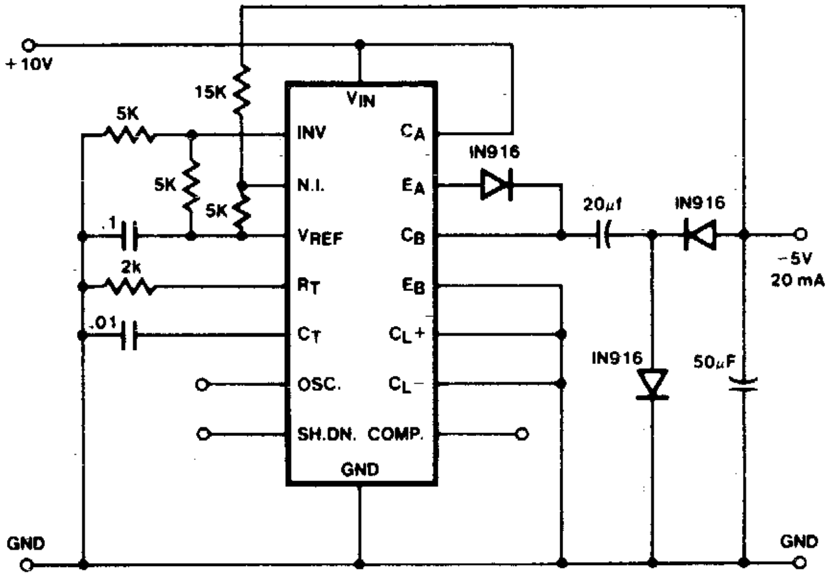
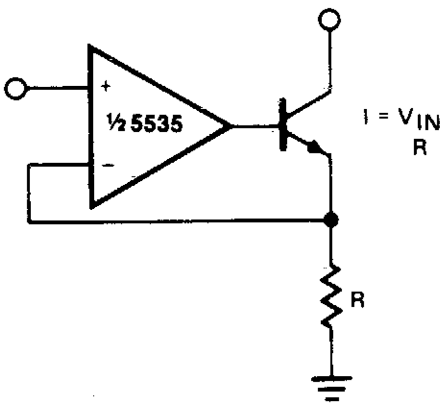


Fig. 17-13

Circuit Notes

The capacitor-diode output circuit is used here as a polarity converter to generate a −5 volt supply from +15 volts. This circuit is useful for an output current of up to 20 mA with no additional boost transistors required. Since the output transistors are current limited, no additional protection is necessary. Also, the lack of an inductor allows the circuit to be stabilized with only the output capacitor.

VOLTAGE-TO-CURRENT CONVERTER

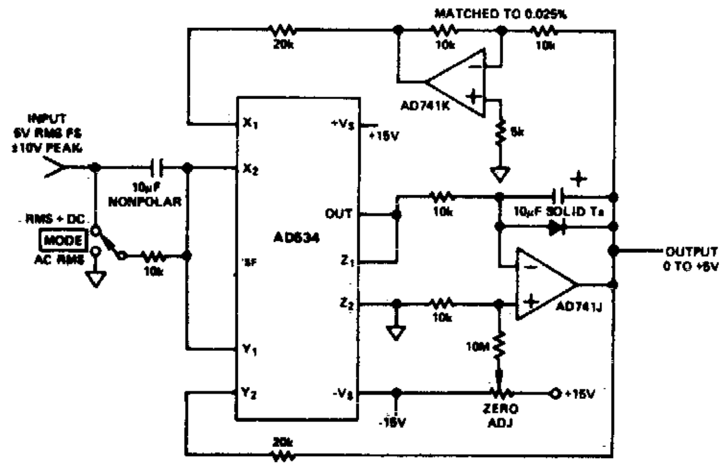


Circuit Notes

The current out is $I_{OUT} \cong V_{IN}/R$. For negative currents, a PNP can be used and, for better accuracy, a Darlington pair can be substituted for the transistor. With careful design, this circuit can be used to control currents of many amps. Unity gain compensation is necessary.

Fig. 17-14

WIDEBAND, HIGH-CREST FACTOR, RMS-TO-DC CONVERTER



CALIBRATION PROCEDURE:

WITH 'MODE' SWITCH IN 'RMS + DC' POSITION, APPLY AN INPUT OF +1.00VDC. ADJUST ZERO UNTIL OUTPUT READS SAME AS INPUT. CHECK FOR INPUTS OF ±10V; OUTPUT SHOULD BE WITHIN ±0.05% (5mV).

ACCURACY IS MAINTAINED FROM 60Hz to 100kHz, AND IS TYPICALLY HIGH BY 0.5% AT 1MHz FOR V_{IN} = 4V RMS (SINE, SQUARE OR TRIANGULAR WAVE).

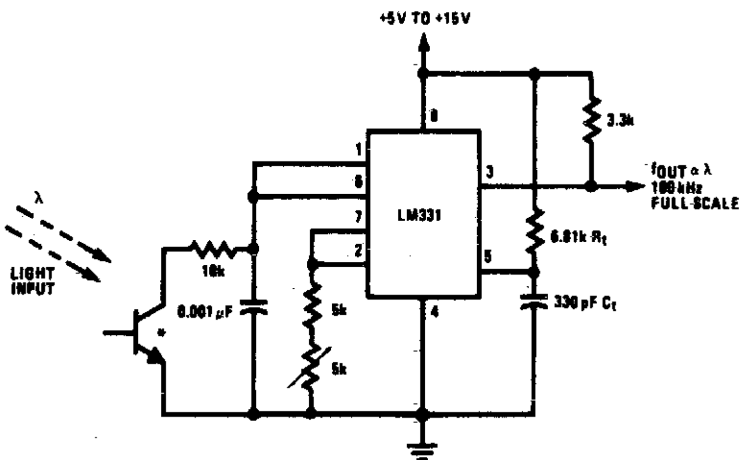
PROVIDED THAT THE PEAK INPUT IS NOT EXCEEDED, CREST-FACTORS UP TO AT LEAST TEN HAVE NO APPRECIABLE EFFECT ON ACCURACY.

INPUT IMPEDANCE IS ABOUT 10kΩ; FOR HIGH (10MΩ) IMPEDANCE, REMOVE MODE SWITCH AND INPUT COUPLING COMPONENTS.

FOR GUARANTEED SPECIFICATIONS THE AD536A AND AD636 IS OFFERED AS A SINGLE PACKAGE RMS-TO-DC CONVERTER.

Fig. 17-15

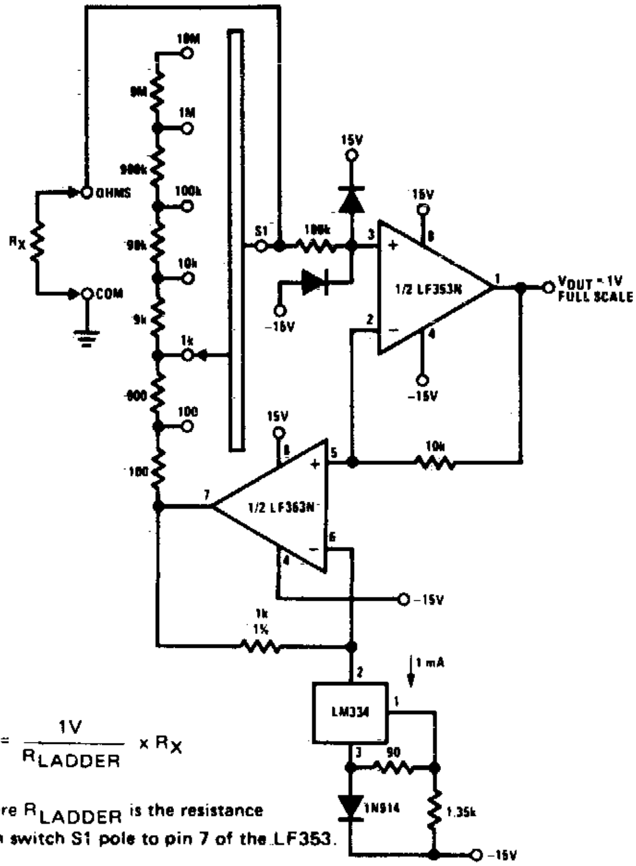
LIGHT INTENSITY-TO-FREQUENCY CONVERTER



*L14F-1, L14G-1 or L14H-1, photo transistor (General Electric Co.) or similar

Fig. 17-16

OHMS-TO-VOLTS CONVERTER



$$V_O = \frac{1V}{R_{LADDER}} \times R_X$$

Where R_{LADDER} is the resistance from switch S1 pole to pin 7 of the LF353.

Fig. 17-17

TEMPERATURE-TO-FREQUENCY CONVERTER

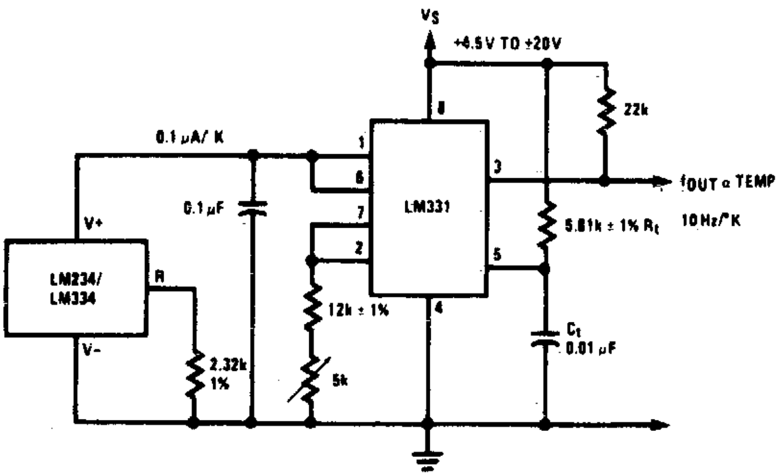
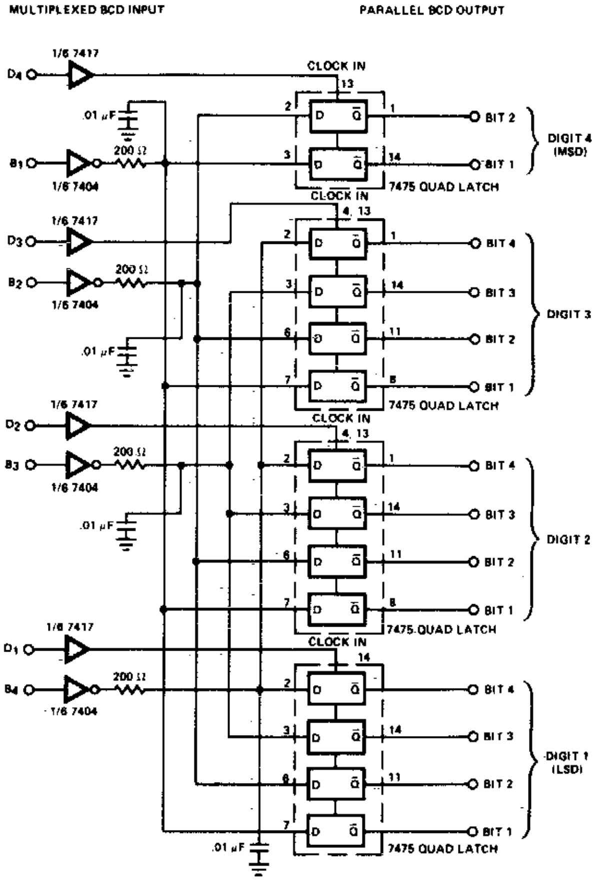


Fig. 17-18

MULTIPLEXED BCD-TO-PARALLEL BCD CONVERTER

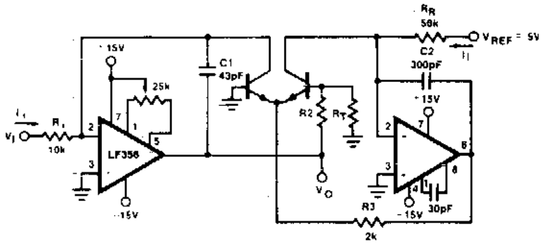


Circuit Notes

Converter consists of four quad bistable latches activated in the proper sequence by the digit strobe output of the LD110. The complemented outputs (Q) of the quad latch set reflects the state of the bit outputs when the digit strobe goes high. It will maintain this state when the digit strobe goes low.

Fig. 17-19

FAST LOGARITHMIC CONVERTER



$$|V_{OUT}| = \left[1 + \frac{R2}{R}\right]\frac{kT}{q} \ln V_I \left[\frac{R_r}{V_{REF} R_I}\right] = \log V_I \frac{1}{R_I I_H}$$

*R2 = 15.7 k, R_1 = 1k, 0.3%/°C (for temperature compensation)

- Dynamic range: 100μA ≤ I_I ≤ 1mA (5 decades, $|V_O|$ = 1V/decades)
- Transient response: 3μs for Δ = decades
- C1, C2, R2, R3: added dynamic compensation
- V_{OS} adjust the LF356 to minimize quiescent error
- R_T: Tel Labs type Q81 + 0.3%/°C.

Fig. 17-20

SINE WAVE-TO-SQUARE WAVE CONVERTER

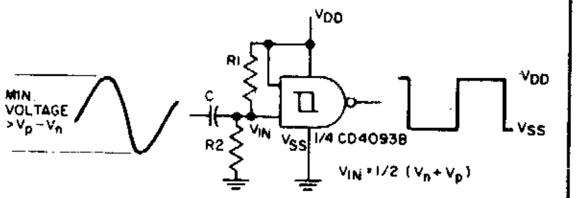


Fig. 17-21

Circuit Notes

The sine input is ac coupled by capacitor C; R1 and R2 bias the input midway between V_n and V_p, the input threshold voltages, to provide a square wave at the output.

TTL-TO-MOS LOGIC CONVERTER

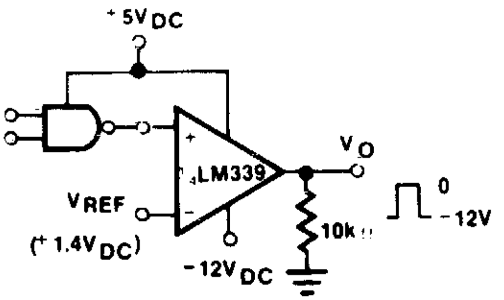


Fig. 17-23

SELF OSCILLATING FLYBACK CONVERTER

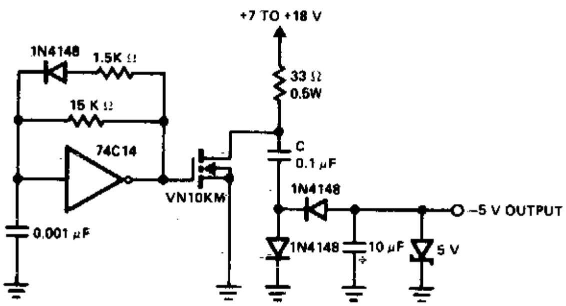


Fig. 17-22

Circuit Notes

A low-power converter suitable for deriving a higher voltage from a main system rail in an on-board application. With the transformer shown, the operating frequency is 250 kHz. Z1 serves as a dissipative voltage regulator for the output and also clips the drain voltage to a level below the rated VMOS breakdown voltage.

PICOAMPERE-TO-VOLTAGE CONVERTER WITH GAIN

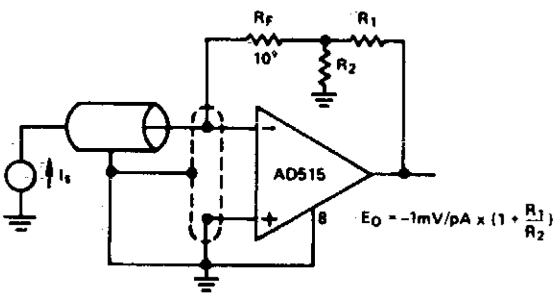


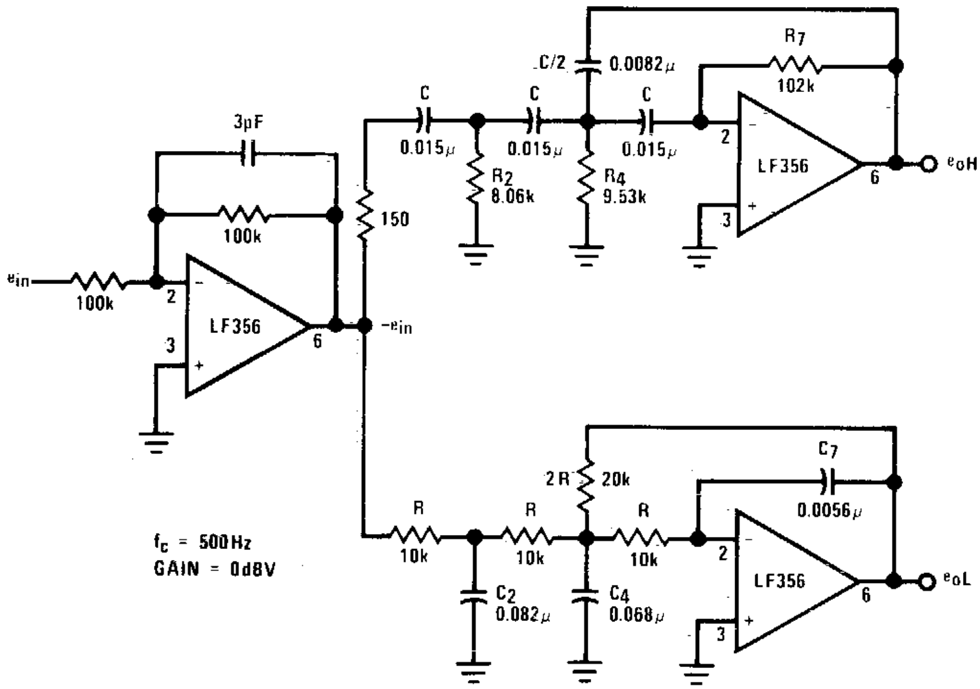
Fig. 17-24

18

Crossover Networks

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ACTIVE CROSSOVER NETWORK



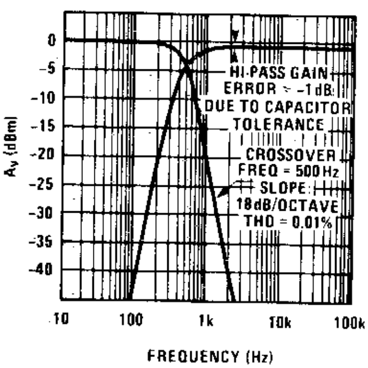


Fig. 18-1

ASYMMETRICAL THIRD ORDER BUTTERWORTH ACTIVE CROSSOVER NETWORK

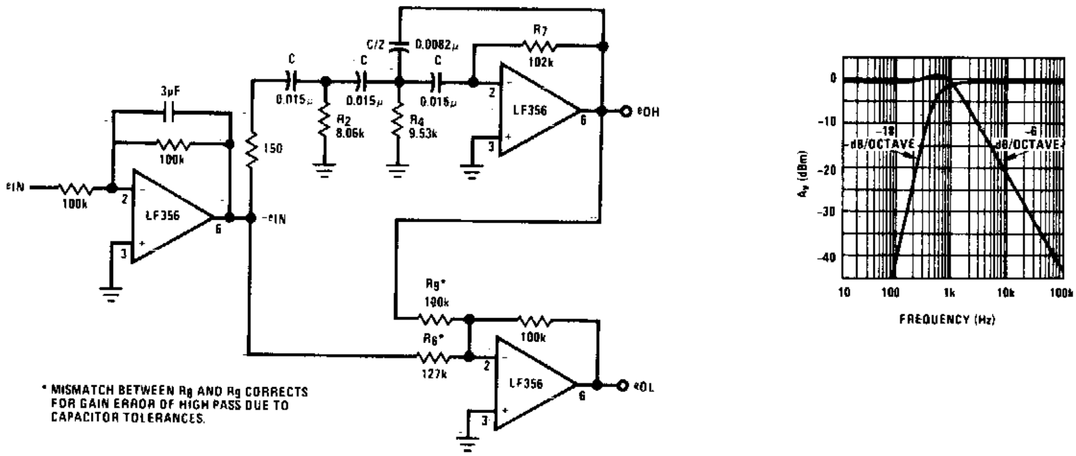
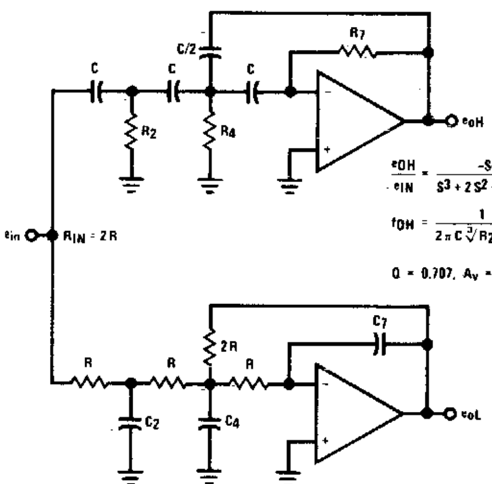


Fig. 18-2

THIRD ORDER BUTTERWORTH CROSSOVER NETWORK



$$\frac{e_{OH}}{e_{IN}} = \frac{-S^3}{S^3 + 2S^2 + 2S + 1}$$

$$f_{OH} = \frac{1}{2\pi C \sqrt[3]{R_2 R_4 R_7}}$$

$Q = 0.707,\ A_V = -1$

Fig. 18-3

$$C_2 = \frac{2.4553}{2\pi f_{OL} R} \qquad R_2 = \frac{0.4074}{2\pi f_{OH} C} \qquad \frac{e_{OL}}{e_{IN}} = \frac{-1}{S^3 + 2S^2 + 2S + 1}$$

$$C_4 = \frac{2.1089}{2\pi f_{OL} R} \qquad R_4 = \frac{0.4742}{2\pi f_{OH} C} \qquad f_{OL} = \frac{1}{2\pi R \sqrt[3]{C_2 C_4 C_7}}$$

$$C_7 = \frac{0.1931}{2\pi f_{OL} R} \qquad R_7 = \frac{5.1766}{2\pi f_{OH} C} \qquad Q = 0.707,\ A_V = -1$$

19

Crystal Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

High Frequency Crystal Oscillator
Overtone Crystal Oscillator
Overtone Crystal Oscillator
TTL Oscillator for 1 MHz-10 MHz
Crystal Checker
96 MHz Crystal Oscillator
Simple TTL Crystal Oscillator
Crystal Oscillator
Overtone Crystal Oscillator
Schmitt Trigger Crystal Oscillator
50 MHz-150 MHz Overtone Oscillator
Fifth Overtone Oscillator
Crystal Controlled Butler Oscillator
Overtone Oscillator with Crystal Switching
Crystal Oscillator
Crystal Oscillator/Doubler
Low Frequency Crystal Oscillator
Crystal Oscillator
100 kHz Crystal Calibrator
Third Overtone Crystal Oscillator
Crystal Checker
CMOS Crystal Oscillator
Temperature-Compensated Crystal Oscillator
Crystal Controlled Transistor Oscillator
Pierce Harmonic Oscillator
Colpitts Harmonic Oscillator
International Crystal OF-1 LO Oscillator
Butler Emitter Follower Oscillator
Colpitts Harmonic Oscillator
Butler Emitter Follower Oscillator
Butler Common Base Oscillator
Pierce Harmonic Oscillator
Tube Type Crystal Oscillator
Precision Clock Generator
Miller Oscillator
Butler Emitter Follower Oscillator
Colpitts Oscillator
Crystal-Controlled Oscillator
Pierce Oscillator
Butler Aperiodic Oscillator
Parallel-mode Aperiodic Crystal Oscillator
International Crystal OF-1 HI Oscillator
Standard Crystal Oscillator for 1 MHz
TTL-Compatible Crystal Oscillator
Crystal Controlled Sine Wave Oscillator
Crystal Oscillator
Stable Low Frequency Crystal Oscillator
JFET Pierce Crystal Oscillator
CMOS Oscillator
Pierce Harmonic Oscillator

HIGH FREQUENCY CRYSTAL OSCILLATOR

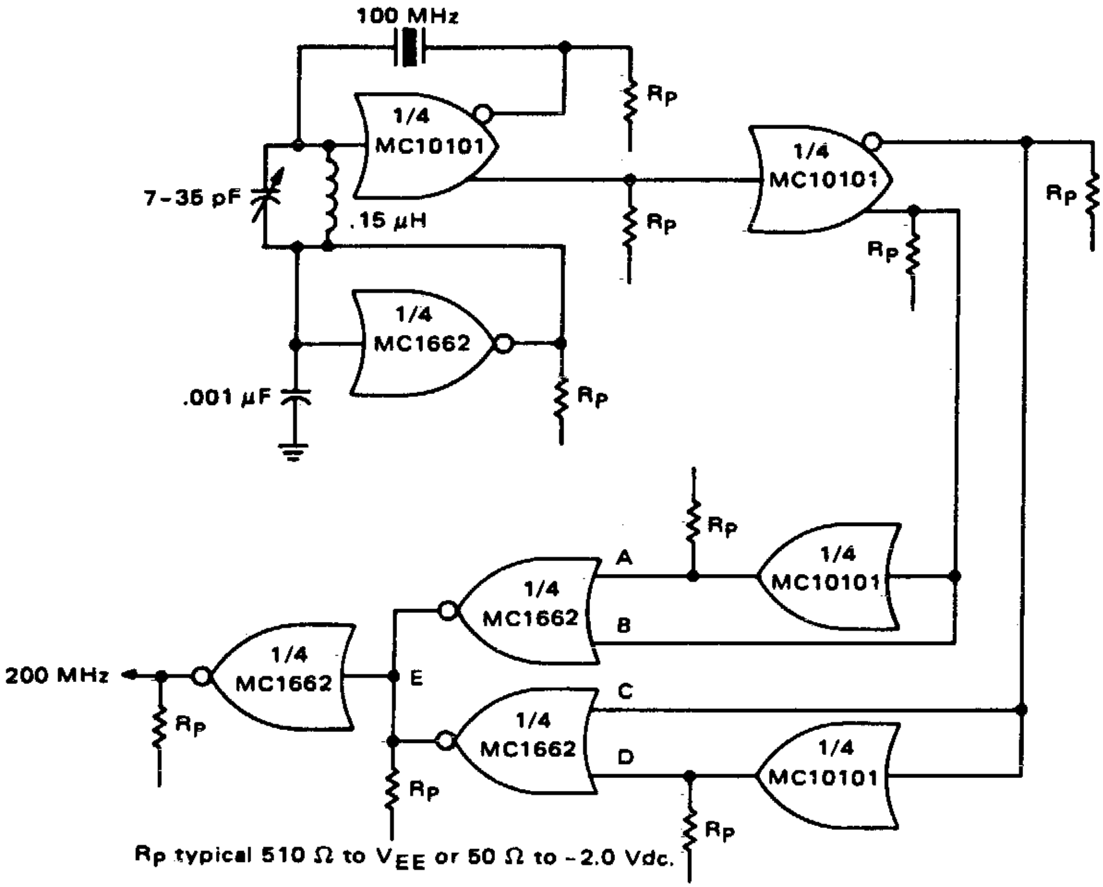


Fig. 19-1

Circuit Notes

One section of the MC10101 is connected as a 100 MHz crystal oscillator with the crystal in series with the feedback loop. The LC tank circuit tunes the 100 MHz harmonic of the crystal and may be used to calibrate the circuit to the exact frequency. A second section of the MC10101 buffers the crystal oscillator and gives complementary 100 MHz signals. The frequency doubler consists of two MC10101 gates as phase shifters and two MC1662 NOR gates. For a 50% duty cycle at the output, the delay to the true and complement 100 MHz signals should be 90°. This may be built precisely with 2.5 ns delay lines for the 200 MHz output or approximated by the two MC10101 gates as shown.

OVERTONE CRYSTAL OSCILLATOR

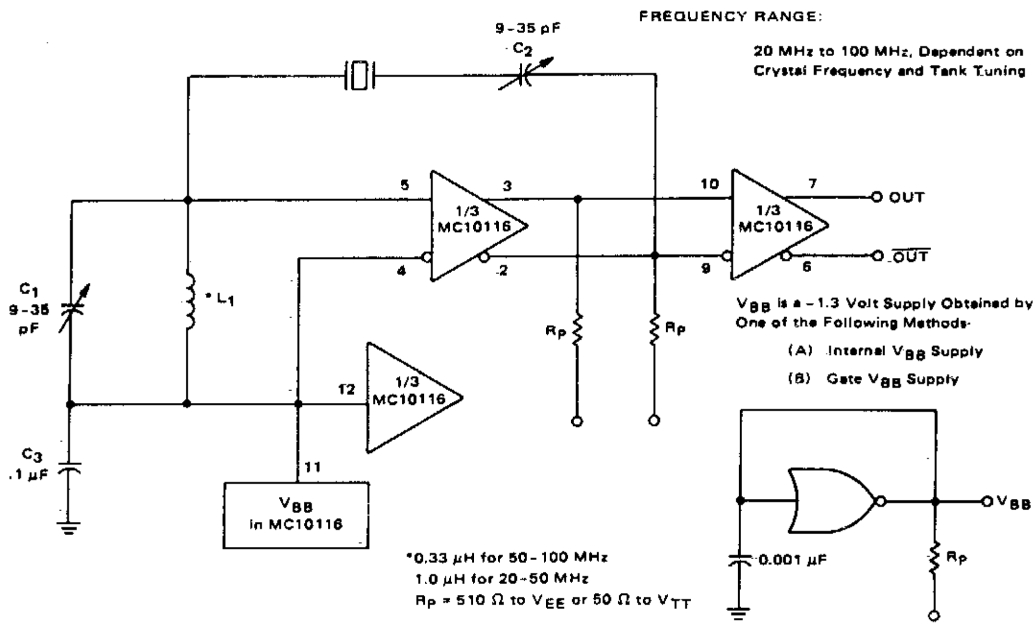


Fig. 19-2

Circuit Notes

This circuit employs an adjustable resonant tank circuit which insures operation at the desired crystal overtone. C1 and L1 form the resonant tank circuit, which with the values specified as a resonant frequency adjustable from approximately 50 MHz to 100 MHz. Overtone operation is accomplished by adjusting the tank circuit frequency at or near the desired frequency. The tank circuit exhibits a low impedance shunt to off-frequency oscillations and a high impedance to the desired frequency, allowing feedback from the output. Operation in this manner guarantees that the oscillator will always start at the correct overtone.

OVERTONE CRYSTAL OSCILLATOR

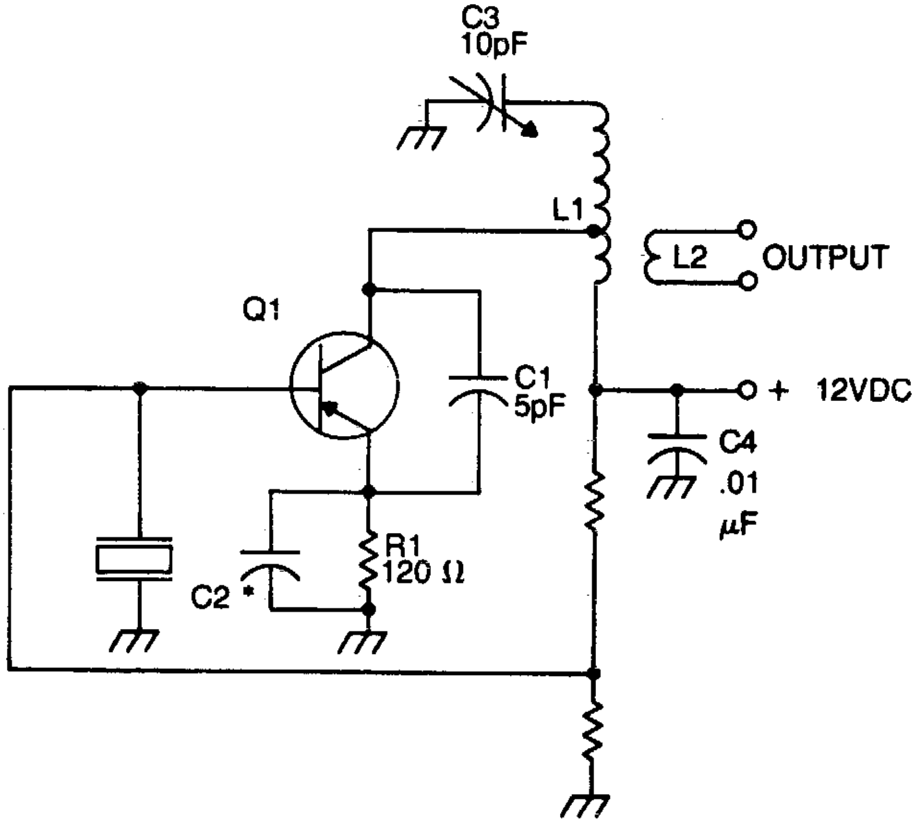


Fig. 19-3

Circuit Notes

The crystal element in this circuit is connected directly between the base and ground. Capacitor C1 is used to improve the feedback due to the internal capacitances of the transistor. This capacitor should be mounted as close as possible to the case of the transistor. The LC tank circuit in the collector of the transistor is tuned to the overtone frequency of the crystal. The emitter resistor capacitor must have a capacitive reactance of approximately 90 ohms at the frequency of operation. The tap on inductor L1 is used to match the impedance of the collector of the transistor. In most cases, the optimum placement of this tap is approximately one-third from the cold end of the coil. The placement of this tap is a trade-off between stability and maximum power output. The output signal is taken from a link coupling coil, L2, and operates by transformer action.

CRYSTAL CHECKER

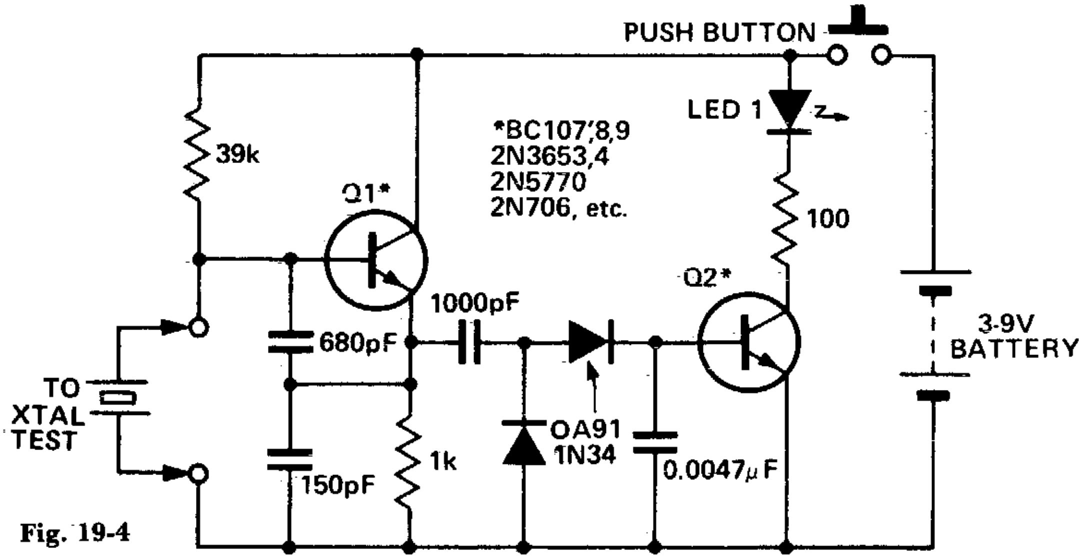


Fig. 19-4

Circuit Notes

Use this circuit for checking fundamental HF crystals on a 'Go-No-Go' basis. An untuned Colpitts oscillator drives a voltage multiplier rectifier and a current amplifier. If the crystal oscillates, Q2 conducts and the LED lights. A3 or 6V, 40mA bulb could be substituted for the LED.

TTL OSCILLATOR FOR 1 MHz-10 MHz

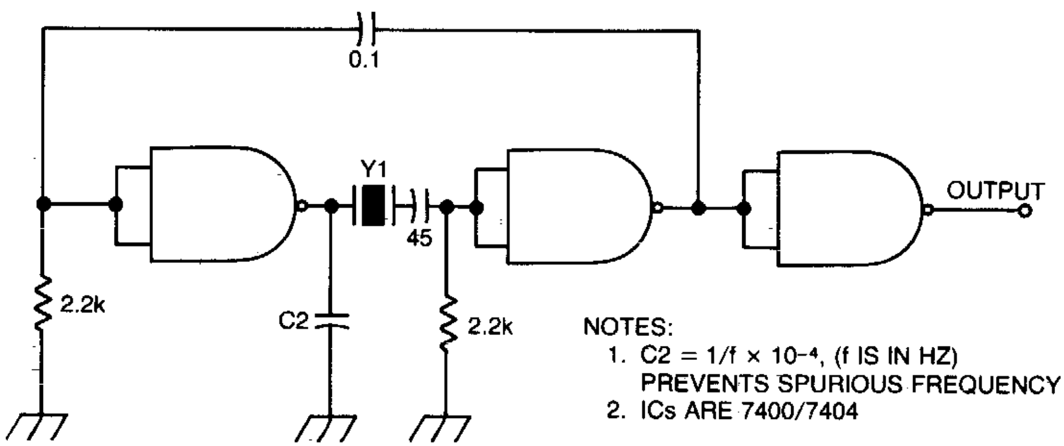
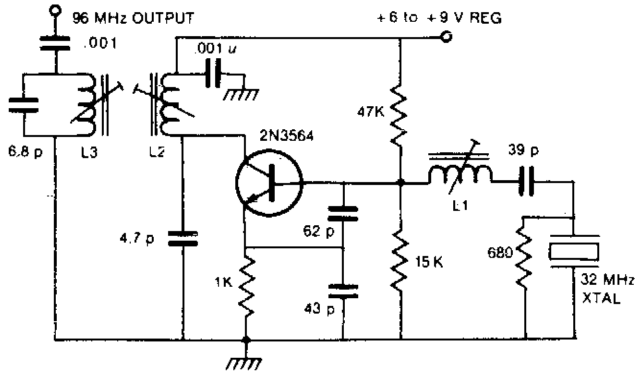


Fig. 19-5

96 MHz CRYSTAL OSCILLATOR



L1, 4 mm former, F29 slug (Neosid AZ assembly) 30 turns .4 mm enamel wire.
L2,L3 7300 CAN TWO 722/1 FORMERS F29 SLUGS (Neosid double assembly) 12 turns .63 mm enamel wire.

Fig. 19-6

Circuit Notes

By using a crystal between 27.5 and 33 MHz, the 3rd harmonic will deliver between 82.5 and 99 MHz.

SIMPLE TTL CRYSTAL OSCILLATOR

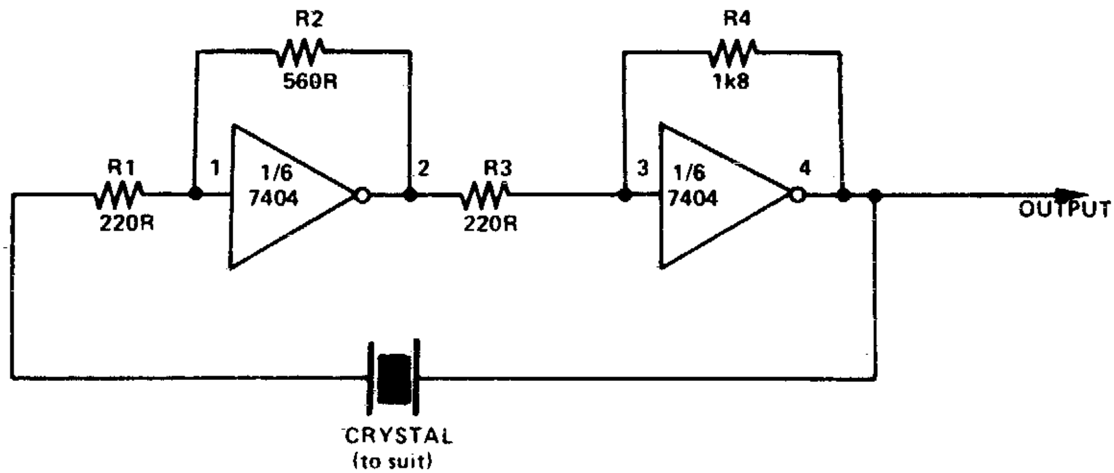


Fig. 19-7

Circuit Notes

This simple and cheap crystal oscillator comprises one third of a 7404, four resistors and a crystal. The inverters are biased into their linear regions by R1 to R4, and the crystal provides the feedback. Oscillation can only occur at the crystals fundamental frequency.

CRYSTAL OSCILLATOR

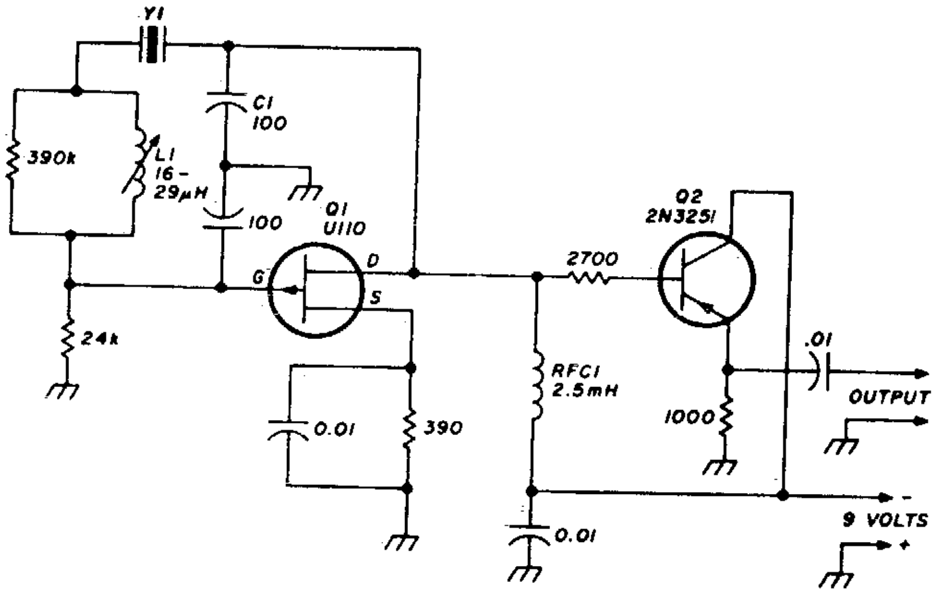


Fig. 19-8

Circuit Notes

Stable VXO using 6- or 8-MHz crystals uses a capacitor and an inductor to achieve frequency pulling on either side of series resonance.

OVERTONE CRYSTAL OSCILLATOR

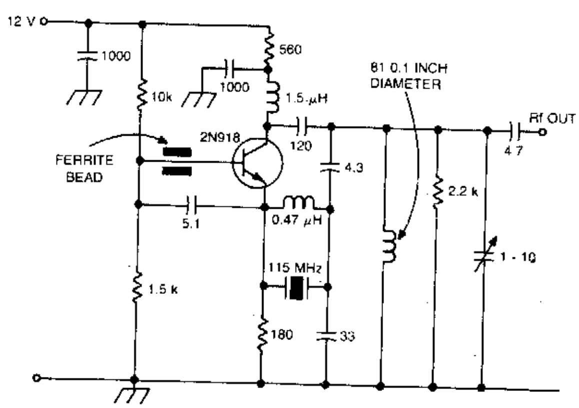


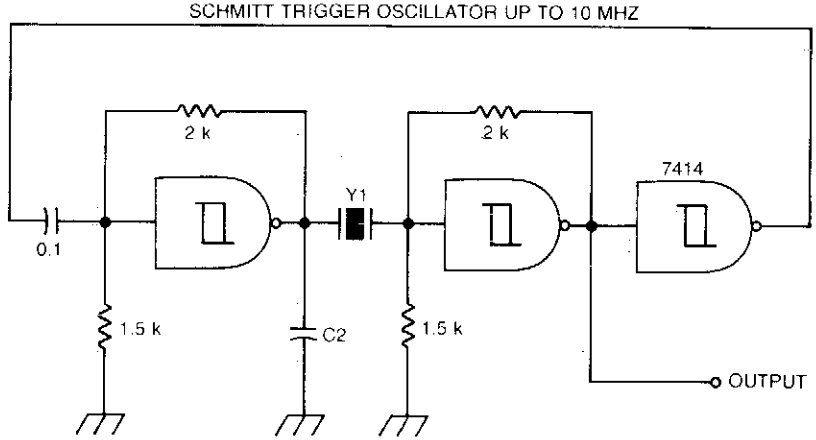
Fig. 19-9

Circuit Notes

This design is for high reliability over a wide temperature range using fifth and seventh overtone crystals. The inductor in parallel with the crystal causes antiresonance of crystal C_0 to minimize loading. This technique is commonly used with overtone crystals.

SCHMITT TRIGGER CRYSTAL OSCILLATOR

SCHMITT TRIGGER OSCILLATOR UP TO 10 MHZ



NOTE: $C2 = 1/f \times 10^{-4}$ (f IS IN HZ) - PREVENTS SPURIOUS FREQUENCY

Fig. 19-10

Circuit Notes

A Schmitt trigger provides good squaring of the output, sometimes eliminating the need for an extra output stage.

50 MHz-150 MHz OVERTONE OSCILLATOR

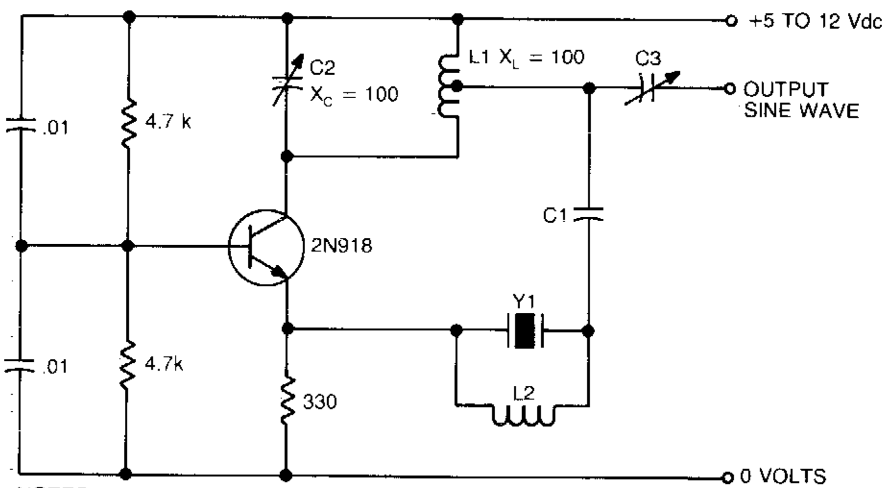


Fig. 19-11

NOTES:

1. Y1 IS AT CUT OVERTONE CRYSTAL.
2. TUNE L1 AND C2 TO OPERATING FREQUENCY.
3. L2 AND SHUNT CAPACITANCE, CO. OF CRYSTAL (APPROXIMATELY 6pF) SHOULD RESONATE TO OSCILLATOR OUTPUT FREQUENCY (L2 = .5 μH AT 90 MHZ). THIS IS NECESSARY TO TUNE OUT EFFECT OF CO.
4. C3 IS VARIED TO MATCH OUTPUT.

FIFTH-OVERTONE OSCILLATOR

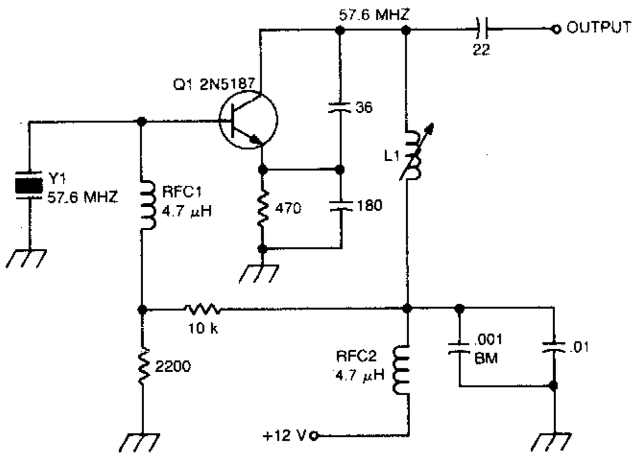


Fig. 19-12

Circuit Notes

This circuit isolates the crystal from the dc base supply with an rf choke for better starting characteristics.

CRYSTAL CONTROLLED BUTLER OSCILLATOR

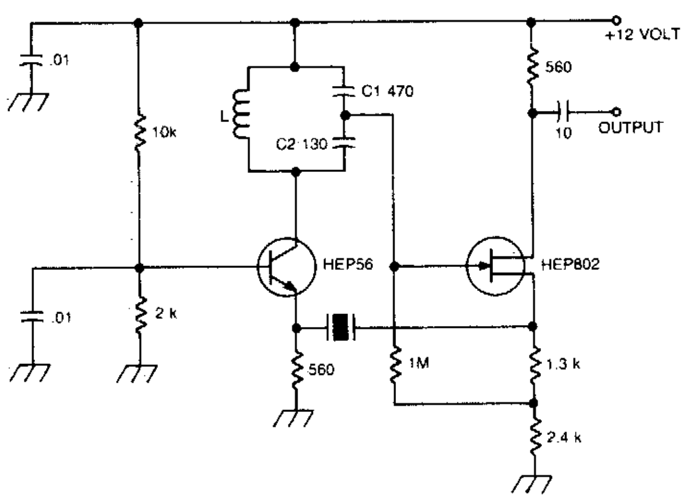
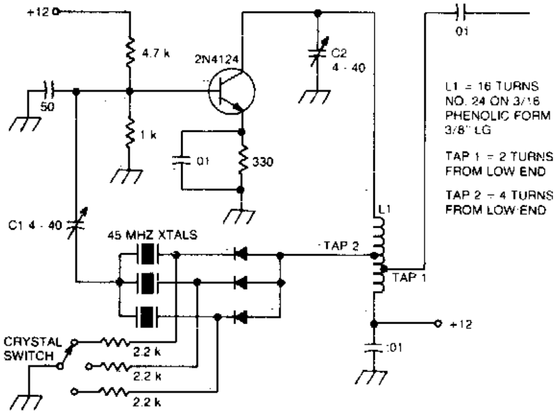


Fig. 19-13

Circuit Notes

A typical Butler oscillator (20-100 MHz) uses an FET in the second stage; the circuit is not reliable with two bipolars. Sometimes two FETs are used. Frequency is determined by LC values.

OVERTONE OSCILLATOR WITH CRYSTAL SWITCHING

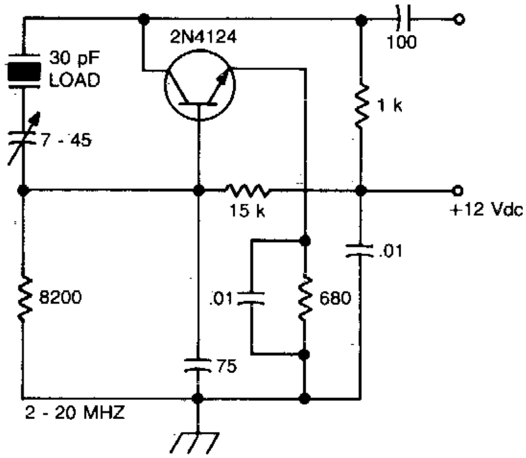


Circuit Notes

The large inductive phase shift of L1 is compensated for by C1. Overtone crystals have very narrow bandwidth; therefore, the trimmer has a smaller effect than for fundamental-mode operation.

Fig. 19-14

CRYSTAL OSCILLATOR



Circuit Notes

The crystal is in a feedback circuit from collector to base. A trimmer capacitor in series shifts the point on the reactance curve where the crystal operates, thus providing a frequency trim. The capacitor has a negative reactance so the crystal is shifted to operate in the positive reactance region.

Fig. 19-15

CRYSTAL OSCILLATOR/DOUBLER

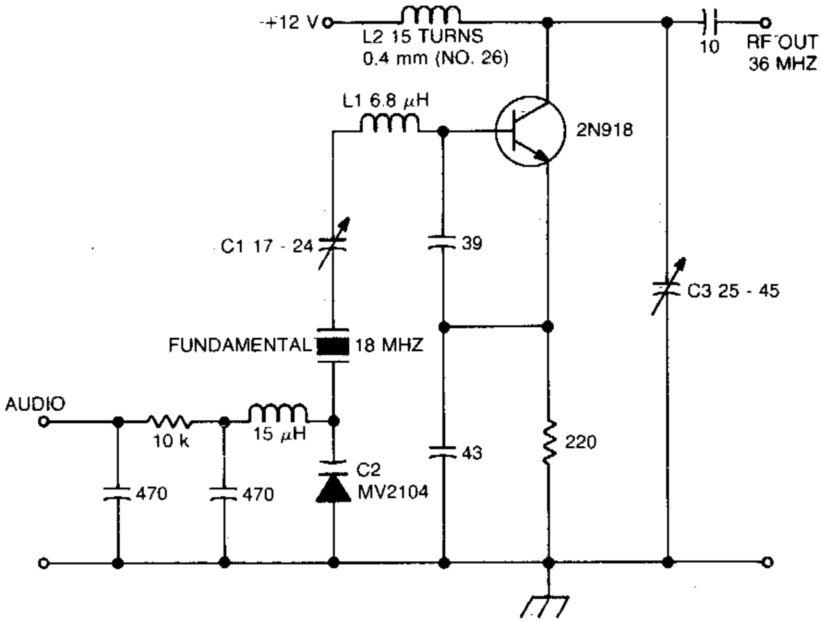


Fig. 19-16

Circuit Notes

The crystal operates into a complex load at series resonance. L1, C1, and C2 balance the crystal at zero reactance. Capacitor C1 fine-tunes the center frequency. Tank circuit L2, C3 doubles the output frequency the circuit operates as an FM oscillator-doubler.

LOW-FREQUENCY CRYSTAL OSCILLATOR

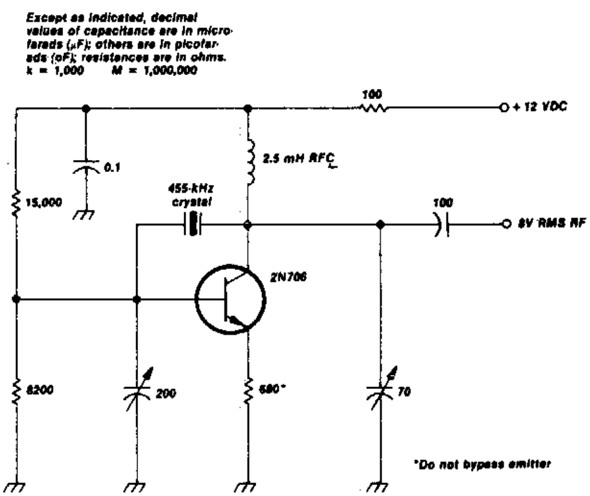


Fig. 19-17

Circuit Notes

This crystal-oscillator circuit uses a 455-kHz crystal.

CRYSTAL OSCILLATOR

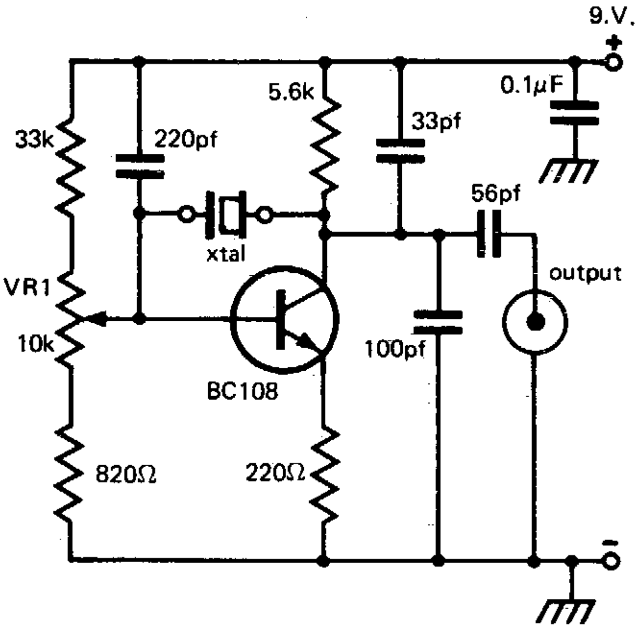


Fig. 19-18

Circuit Notes

This circuit provides reliable oscillation and an output close to one volt peak-to-peak. Power consumption is around 1 mA from a nine volt supply.

100 kHz CRYSTAL CALIBRATOR

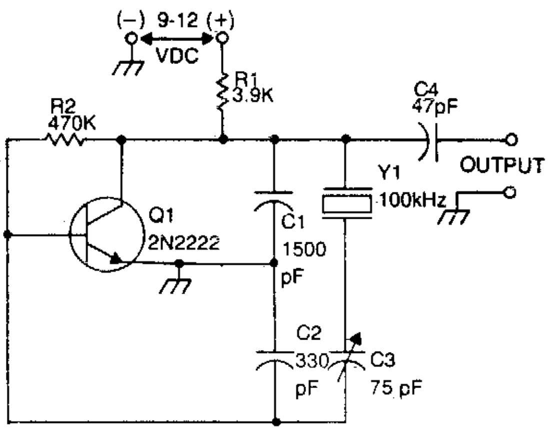


Fig. 19-19

Circuit Notes

This circuit is often used by amateur radio operations, shortwave listeners, and other operators of shortwave receivers to calibrate the dial pointer. The oscillator operates at a fundamental frequency of 100 kHz, and the harmonics are used to locate points on the shortwave dial, provided that the output of the calibrator is coupled to the antenna circuit of the receiver. The crystal shunts the feedback voltage divider, and is in series with a variable capacitor (C3) that is used to set the actual operating frequency of the calibrator.

THIRD-OVERTONE CRYSTAL OSCILLATOR

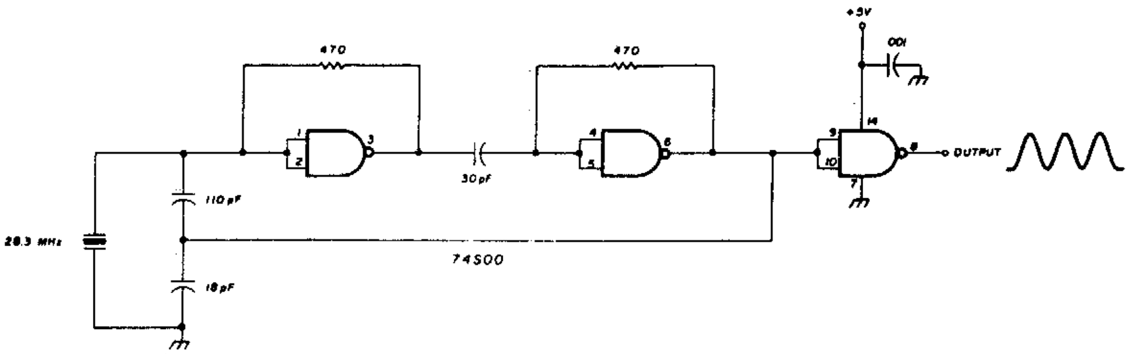
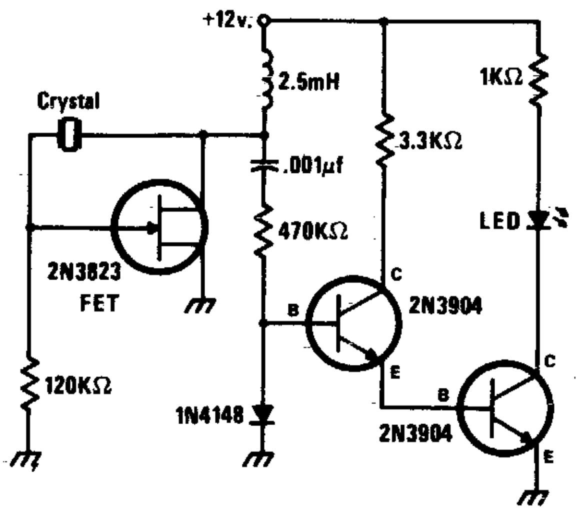


Fig. 19-20

Circuit Notes

This circuit uses a 74S00 Schottky TTL gate; no inductors are required.

CRYSTAL CHECKER



Circuit Notes

This circuit is a simple Pierce oscillator with an LED go/no go display. Checker works best with crystals having fundamental frequencies in the seven to eight megahertz range.

Fig. 19-21

CMOS CRYSTAL OSCILLATOR

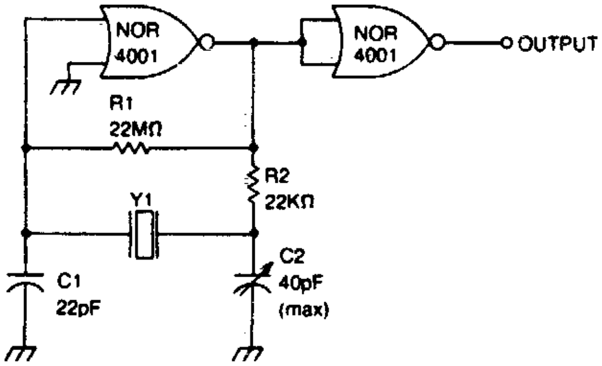


Fig. 19-22

Circuit Notes

This circuit has a frequency range of 0.5 MHz to 2.0 MHz. Frequency can be adjusted to a precise value with trimmer capacitor C2. The second NOR gate serves as an output buffer.

TEMPERATURE-COMPENSATED CRYSTAL OSCILLATOR

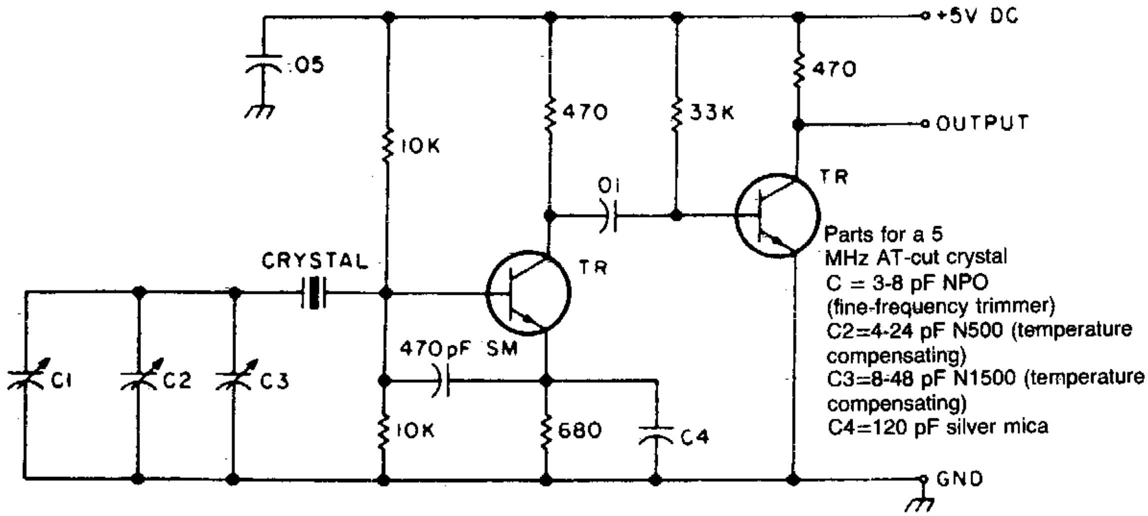


Fig. 19-23

Circuit Notes

Two different negative-coefficient capacitors are blended to produce the desired change in capacitance to counteract or compensate for the decrease in frequency of the "normal" AT-cut characteristics.

CRYSTAL-CONTROLLED, TRANSISTOR OSCILLATOR

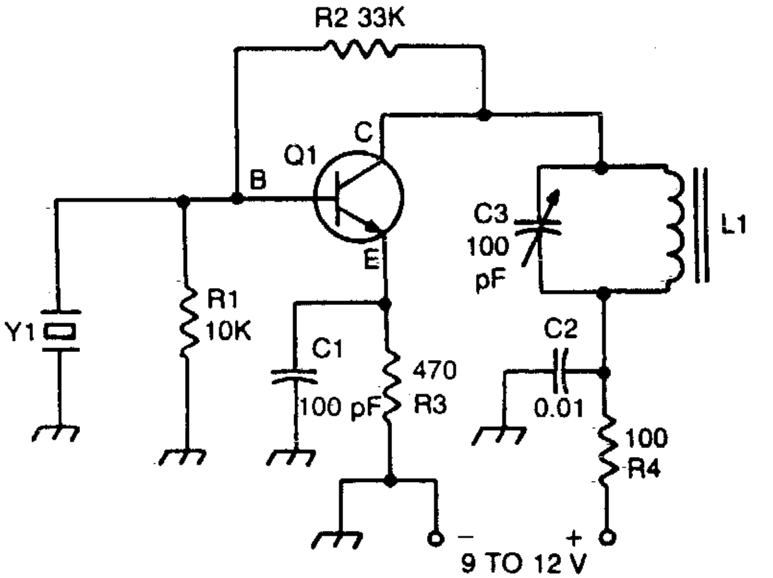


Fig. 19-24

PIERCE HARMONIC OSCILLATOR (20 MHz)

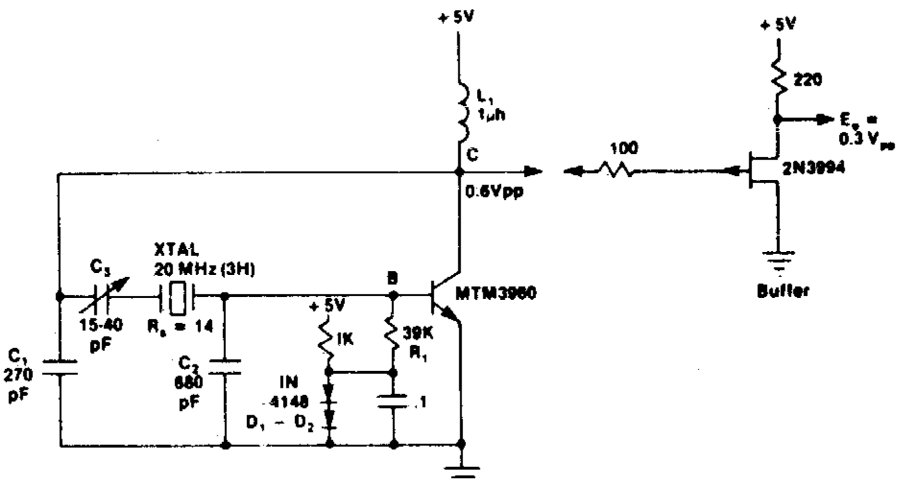


Fig. 19-25

Circuit Notes

This circuit has excellent short term frequency stability because the external load tied across the crystal is mostly capacitive rather than resistive, giving the crystal a high in-circuit Q.

COLPITTS HARMONIC OSCILLATOR (100 MHz)

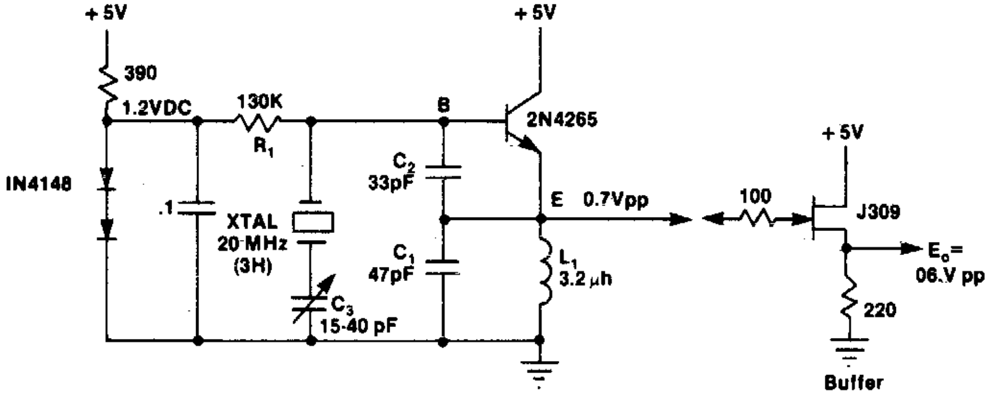
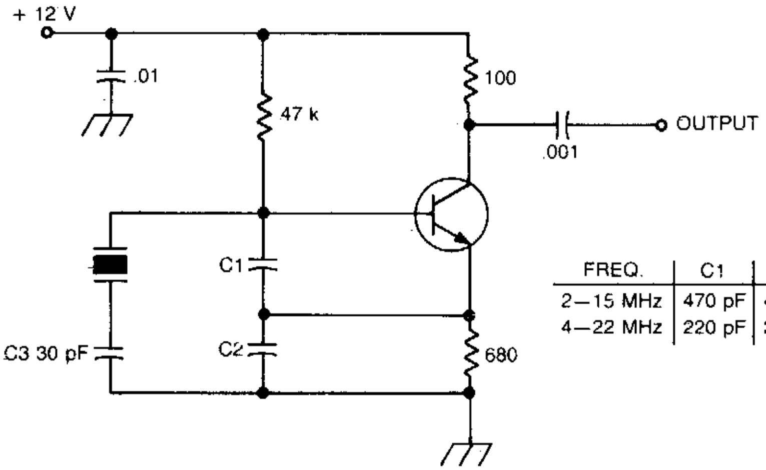


Fig. 19-26

Circuit Notes

L1C1 are selected to be resonant at a frequency below the desired crystal harmonic but above the crystal's next lower odd harmonic. C2 should have a value of 30-70 pF, independent of the oscillation frequency. There is no requirement for any specific ratio of C1/C2, but practical harmonic circuits seem to work best when C1 is approximately 1-3 times the value of C2. Diodes D1-D3 provide a simple regulated bias supply. The resistance of R1 should be as high as possible, as it affects the crystal's in-circuit Q.

INTERNATIONAL CRYSTAL OF-1 LO OSCILLATOR



FREQ.	C1	C2
2—15 MHz	470 pF	470 pF
4—22 MHz	220 pF	220 pF

Fig. 19-27

Circuit Notes

International Crystal OF-1 LO oscillator circuit for fundamental-mode crystals.

BUTLER EMITTER FOLLOWER OSCILLATOR (100 MHz)

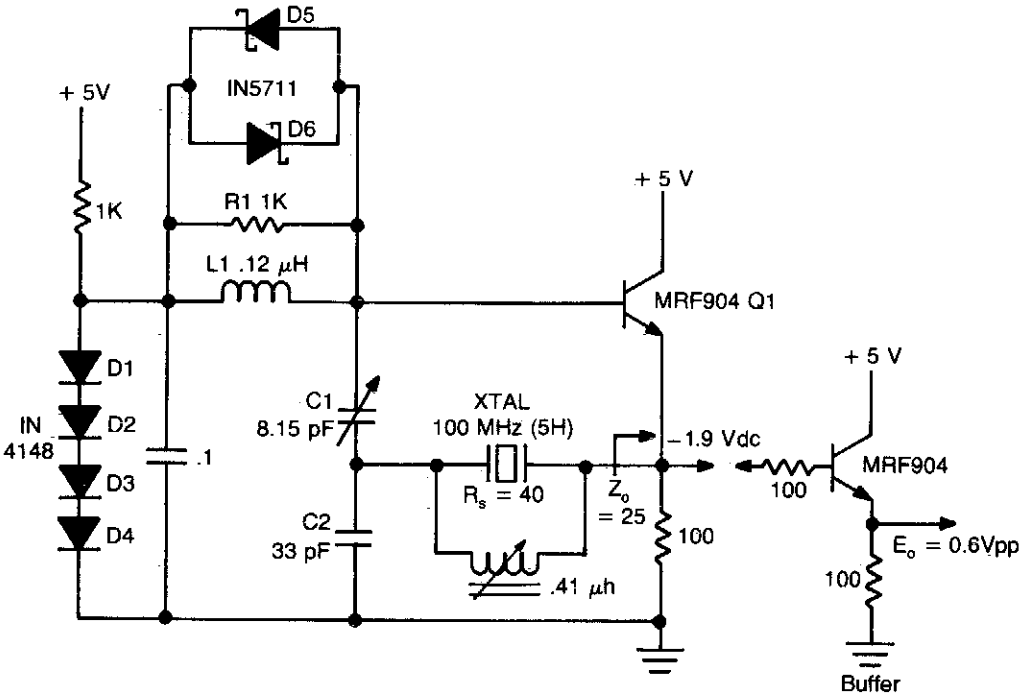
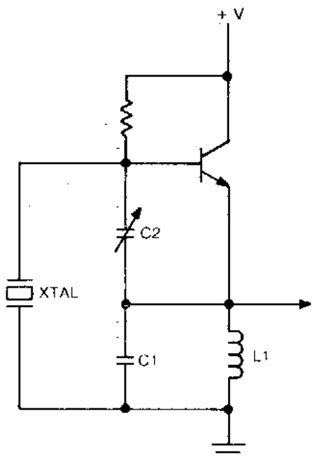


Fig. 19-28

Circuit Notes

This circuit has good performance without any parasitics because emitter follower amplifier has a gain of only one with built-in negative feedback to stabilize its gain.

COLPITTS HARMONIC OSCILLATOR (BASIC CIRCUIT)

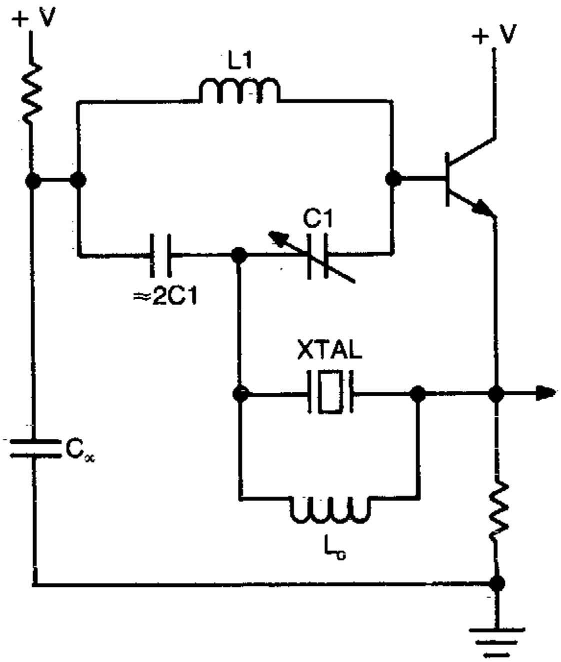


Circuit Notes

This circuit operates 30-200 ppm above series resonance. Physically simple, but analytically complex. It is inexpensive with fair frequency stability.

Fig. 19-29

BUTLER EMITTER FOLLOWER OSCILLATOR (BASIC CIRCUIT)

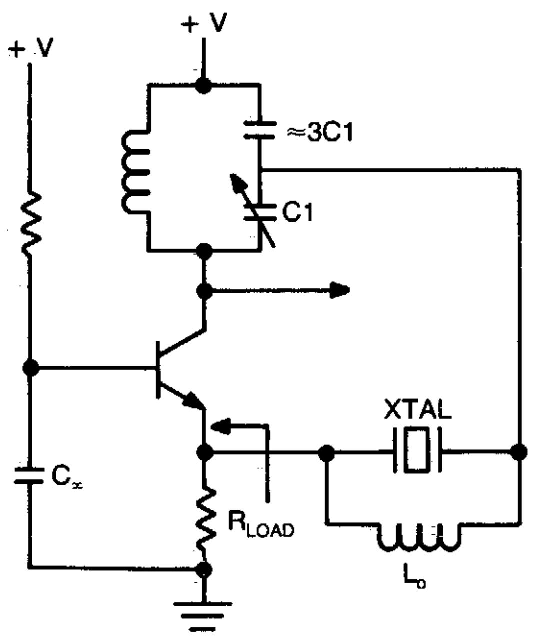


Circuit Notes

This circuit operates at or near series resonance. It is a good circuit design with no parasitics. It is easy to tune with good frequency stability.

Fig. 19-30

BUTLER COMMON BASE OSCILLATOR (BASIC CIRCUIT)

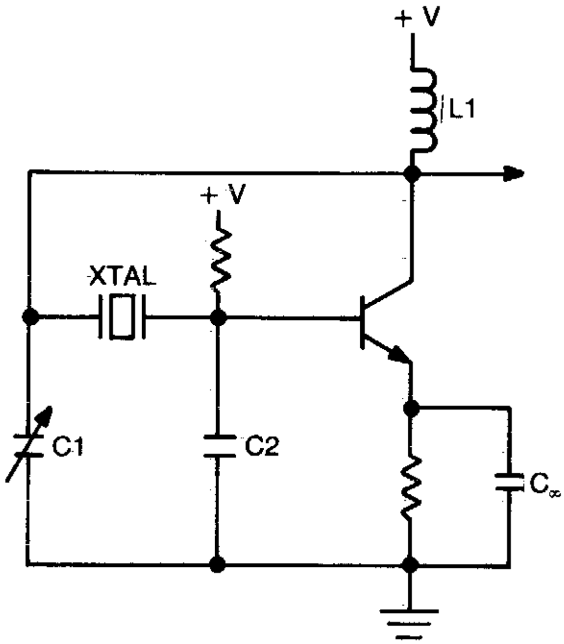


Circuit Notes

This circuit operates at or near series resonance. It has fair to poor circuit design with parasitics, touch to tune, and fair frequency stability.

Fig. 19-31

PIERCE HARMONIC OSCILLATOR (BASIC CIRCUIT)

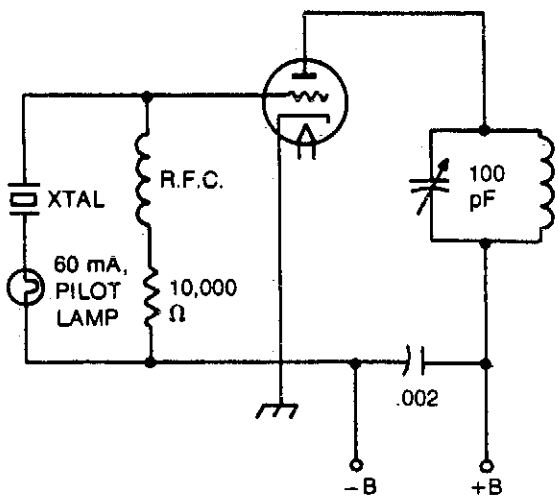


Circuit Notes

This circuit operates 10-40 ppm above series resonance. It is a good circuit design with good to very good frequency stability.

Fig. 19-32

TUBE-TYPE CRYSTAL OSCILLATOR



Circuit Notes

The pilot lamp limits current to prevent damage to the crystal.

Fig. 19-33

PRECISION CLOCK GENERATOR

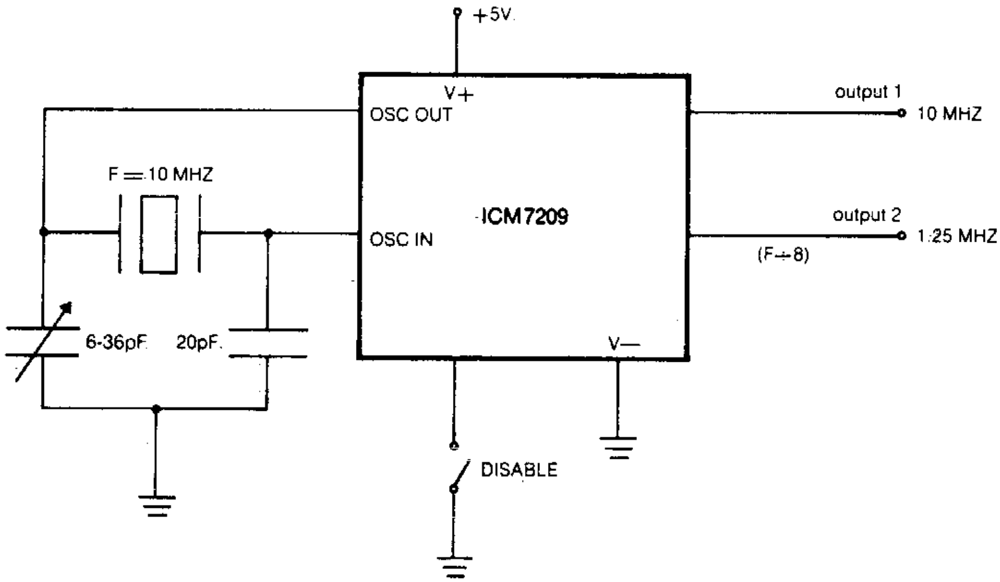
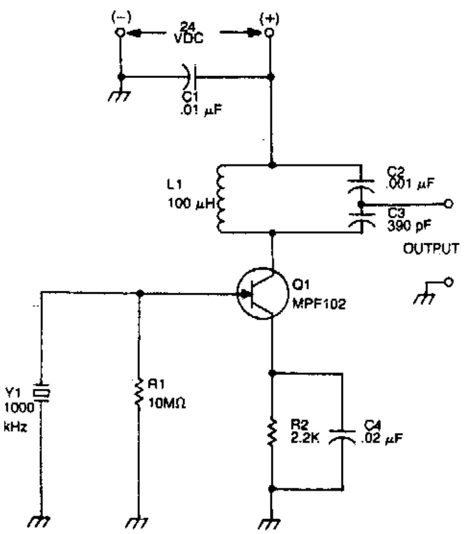


Fig. 19-34

Circuit Notes

The CMOS IC directly drives 5 TTL loads from either of 2 buffered outputs. The device operates to 10 MHz and is bipolar, MOS, and CMOS compatible.

MILLER OSCILLATOR (CRYSTAL CONTROLLED)



Circuit Notes

The drain of the JFET Miller oscillator is tuned to the resonant frequency of the crystal by an LC tank circuit.

Fig. 19-35

BUTLER EMITTER FOLLOWER OSCILLATOR (20 MHz)

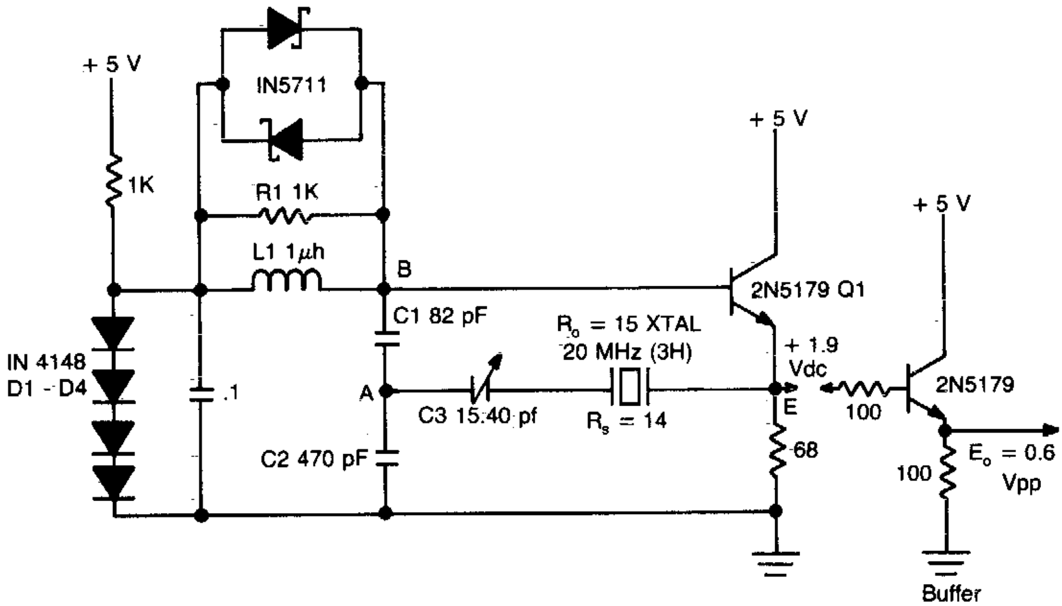
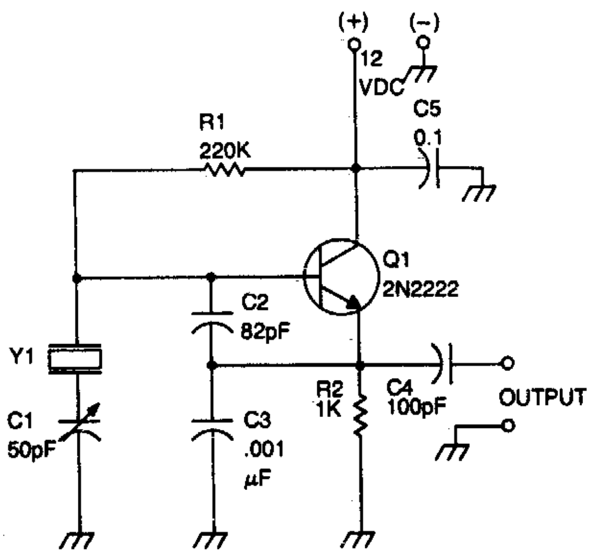


Fig. 19-36

COLPITTS OSCILLATOR



Circuit Notes

This circuit will operate with fundamental-mode crystals in the range of 1 MHz to 20 MHz. Feedback is controlled by capacitor voltage divider C2/C3. The rf voltage across the emitter resistor provides the basic feedback signal.

Fig. 19-37

CRYSTAL-CONTROLLED OSCILLATOR

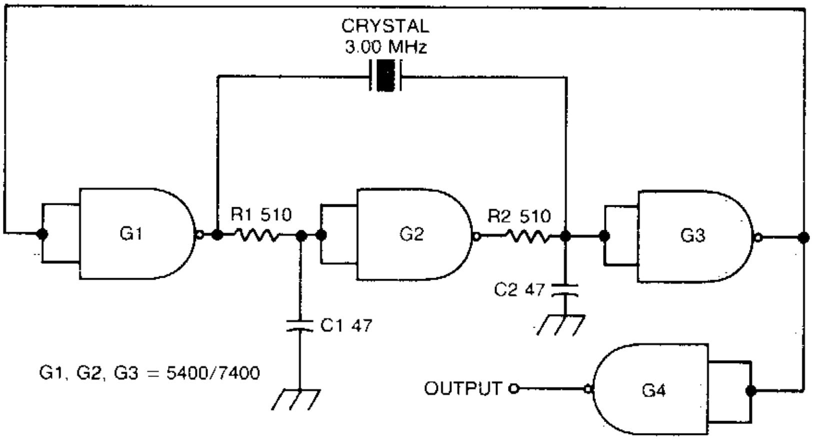


Fig. 19-38

Circuit Notes

This circuit oscillates without the crystal. With the crystal in the circuit, the frequency will be that of the crystal. The circuit has good starting characteristics even with the poorest crystals.

PIERCE OSCILLATOR

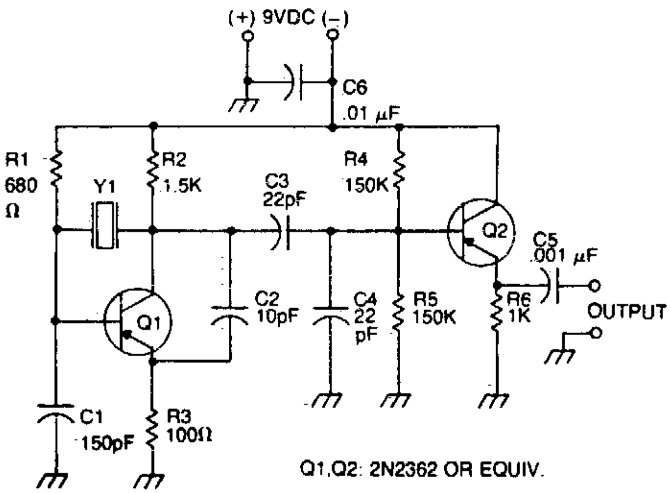


Fig. 19-39

Circuit Notes

The oscillator transistor is Q1, and the crystal is placed between the collector and base. Feedback is improved by the use of the collector-emitter capacitor C2. Transistor Q2 is used as an output buffer.

BUTLER APERIODIC OSCILLATOR

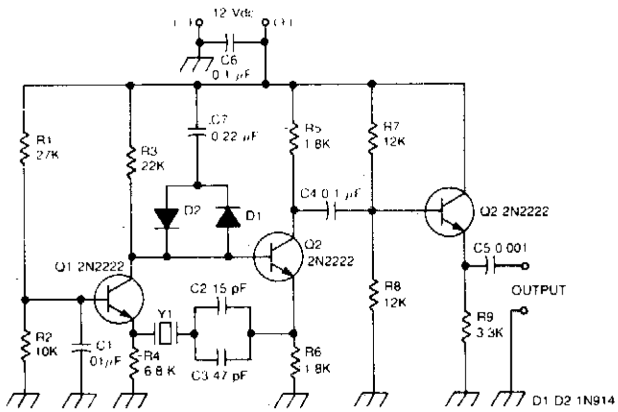


Fig. 19-40

Circuit Notes

This circuit works well in the range of 50 kHz to 500 kHz. Slight component modifications are needed for higher frequency operation. For operation over 3000 kHz, select a transistor that provides moderate gain (in the 60 to 150 range) at the frequency of operation and a gain-bandwidth product of at least 100 MHz.

PARALLEL-MODE APERIODIC CRYSTAL OSCILLATOR

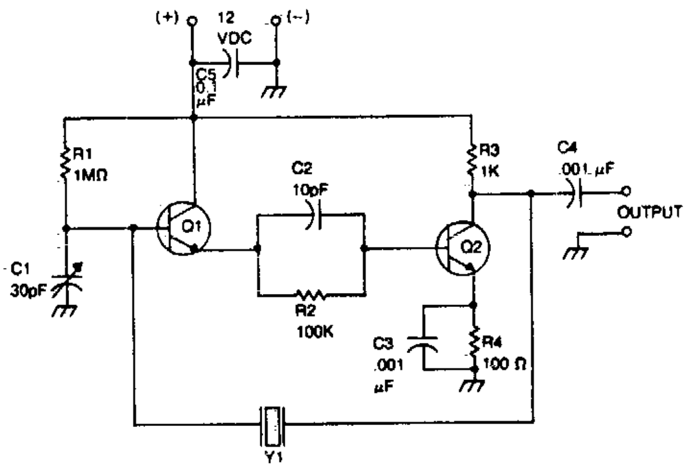
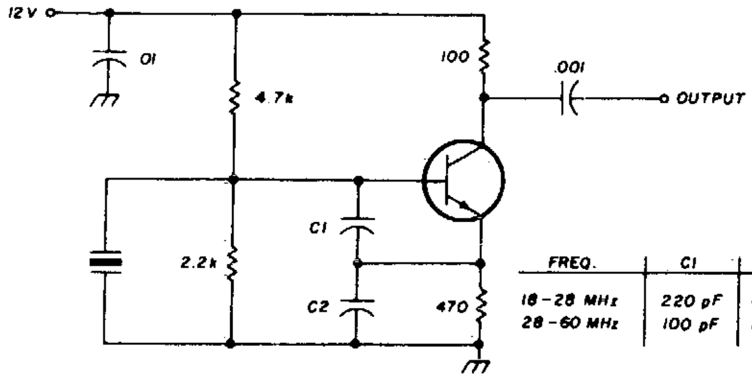


Fig. 19-41

Circuit Notes

The crystal is placed between the collector of the output stage and the base of the input stage. The frequency of oscillation can be set to a precise value with trimmer capacitor C1. The range of operation for this circuit is 500 kHz to 10 MHz. Extend the range downward (100 kHz) by increasing the value of C1 to 75 pF and increasing the value of C2 to 22pF.

INTERNATIONAL CRYSTAL OF-1 HI OSCILLATOR



FREQ.	C1	C2
18–28 MHz	220 pF	47 pF
28–60 MHz	100 pF	18 pF

Fig. 19-42

Circuit Notes

International Crystal OF-1 HI oscillator circuit for third-overtone crystals. The circuit does not require inductors.

STANDARD CRYSTAL OSCILLATOR FOR 1 MHz

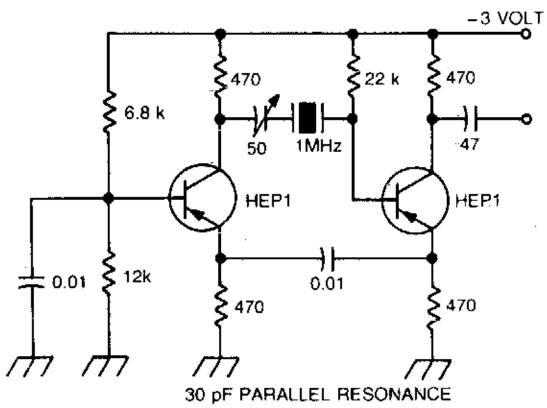


Fig. 19-43

TTL-COMPATIBLE CRYSTAL OSCILLATOR

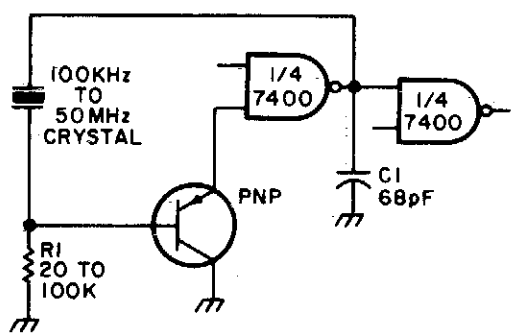


Fig. 19-44

Circuit Notes

Adjust R1 for about 2 volts at the output of the first gate. Adjust C1 for best output.

CRYSTAL CONTROLLED SINE WAVE OSCILLATOR

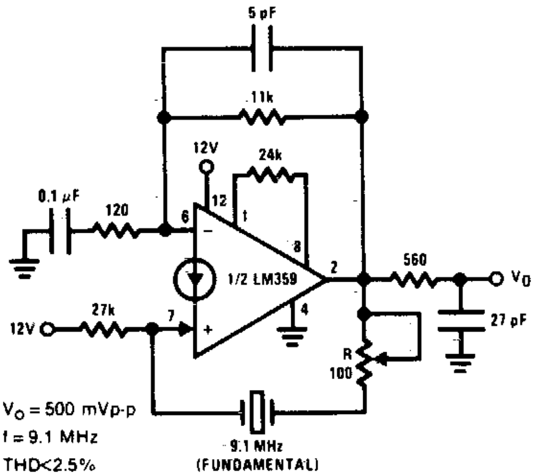


Fig. 19-45

STABLE LOW FREQUENCY CRYSTAL OSCILLATOR

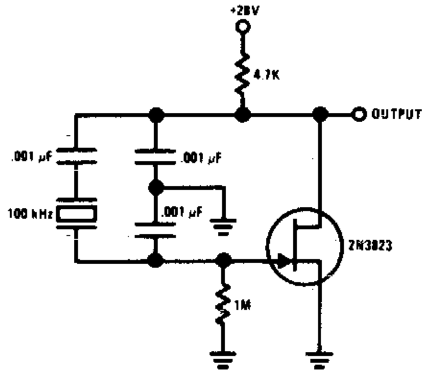


Fig. 19-47

Circuit Notes

This Colpitts-crystal oscillator is ideal for low frequency crystal oscillator circuits. Excellent stability is assured because the 2N3823 JFET circuit loading does not vary with temperature.

CRYSTAL OSCILLATOR

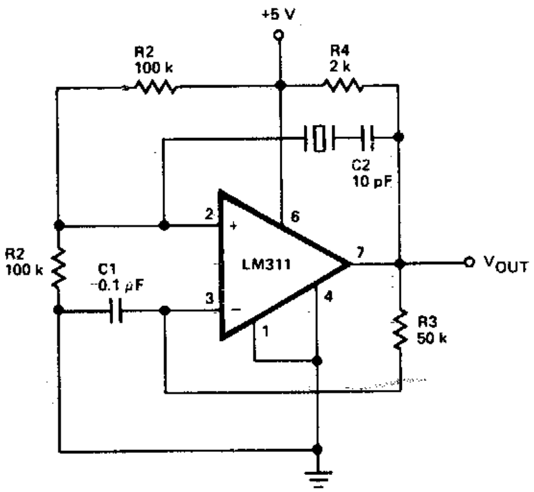


Fig. 19-46

JFET PIERCE CRYSTAL OSCILLATOR

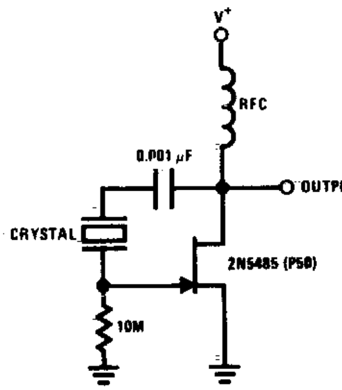


Fig. 19-48

Circuit Notes

The JFET Pierce crystal oscillator allows a wide frequency range of crystals to be used without circuit modification. Since the JFET gate does not load the crystal, good Q is maintained, thus insuring good frequency stability.

CMOS OSCILLATOR-1 MHz-4 MHz

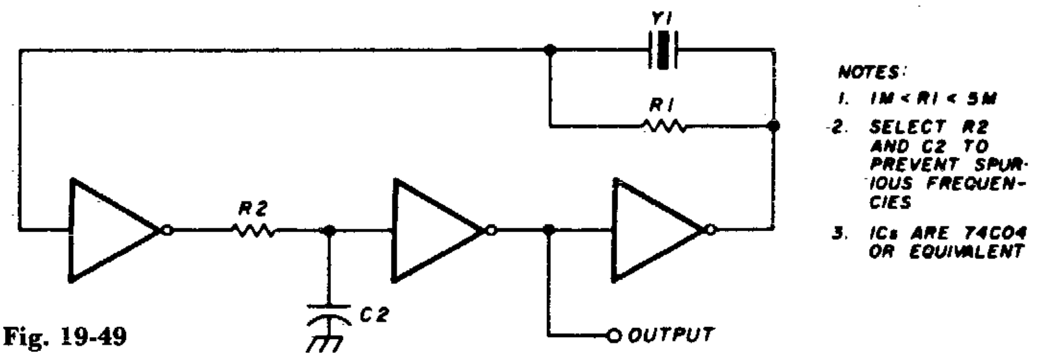


Fig. 19-49

PIERCE HARMONIC OSCILLATOR (100 MHz)

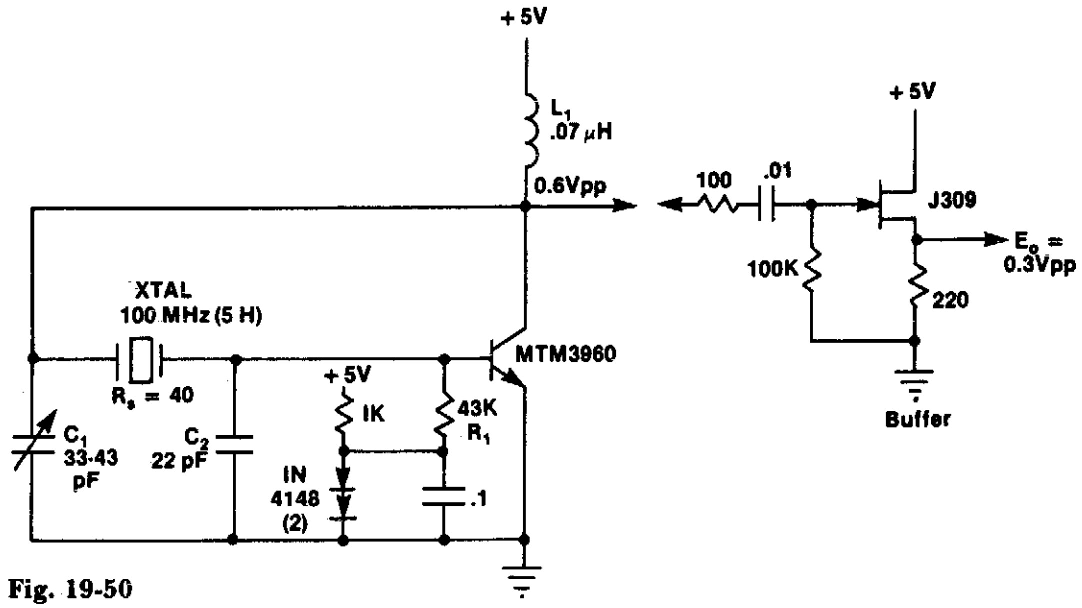


Fig. 19-50

Circuit Notes

The output resistance of the transistor's collector, together with the effective value of C1, provides an RC phase lag of 30-50°. The crystal normally oscillates slightly above series resonance, where it is both resistive and inductive. Above series resonance, the crystal's internal impedance (resistive and inductive) together with C2 provides an RLC phase lag of 130-150°. The transistor inverts the signal, providing a total of 360° of phase shift around the loop. Inductor L1 is selected to resonate with C1 at a frequency between the crystal's desired harmonic and its next lower odd harmonic. Inductor L1 offsets part of the negative reactance of C1 at the oscillation frequency.

20

Current Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AMMETER

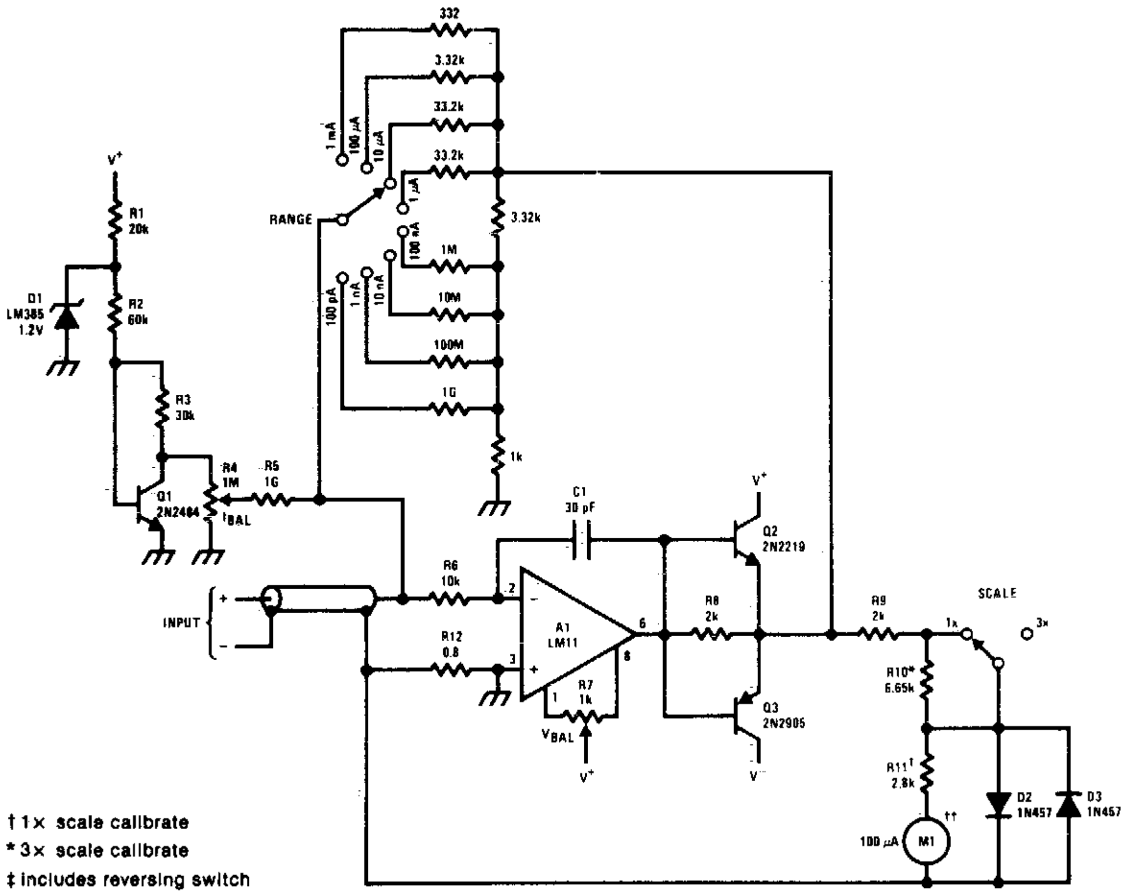


Fig. 20-1

Circuit Notes

Current meter ranges from 100 pA to 3 mA full scale. Voltage across input is 100 μV at lower ranges rising to 3 mV at 3 mA. The buffers on the op amp are to remove ambiguity with high-current overload. The output can also drive a DVM or a DPM.

PICO AMMETER

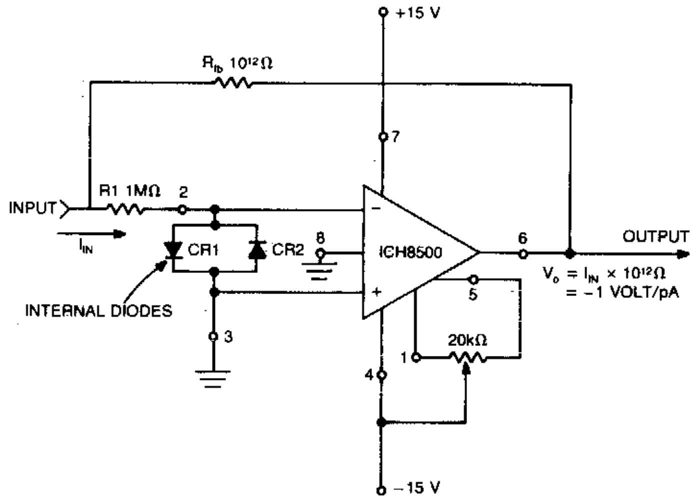
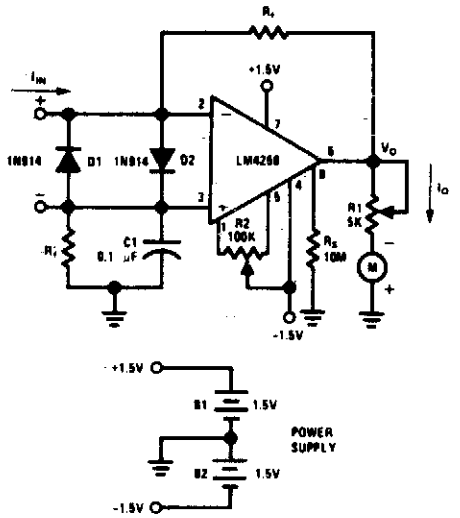


Fig. 20-2

Circuit Notes

A very sensitive pico ammeter (−1 V/pA) employs the amplifier in the inverting or current summing mode. Care must be taken to eliminate stray currents from flowing into the current summing mode. It takes approximately 5 for the circuit to stabilize to within 1% of its final output voltage after a step function of input current has been applied. The internal diodes CR1 and CR2 together with external resistor R1 to protect the input stage of the amplifier from voltage transients.

NANO AMMETER



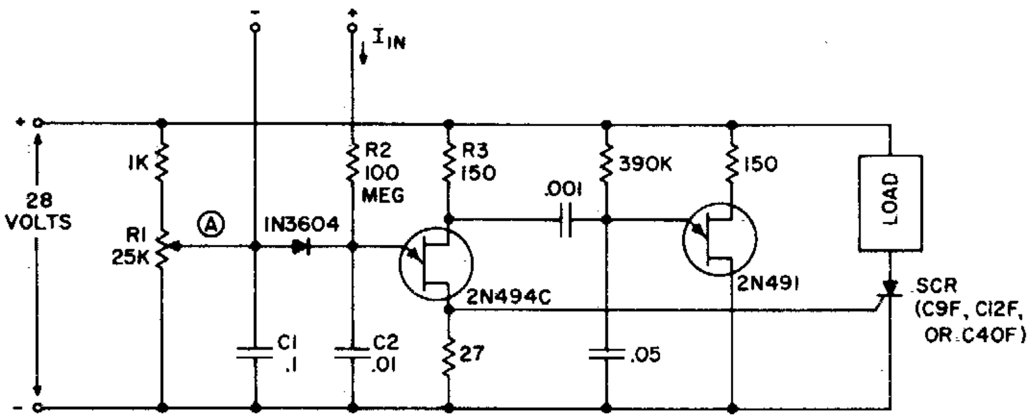
Resistance Values for DC Nano and Micro Ammeter

I FULL SCALE	R_f (Ω)	R_i (Ω)
100 nA	1.5M	1.5M
500 nA	300k	300k
1 µA	300k	0
5 µA	60k	0
10 µA	30k	0
50 µA	6k	0
100 µA	3k	0

The complete meter amplifier is a differential current-to-voltage converter with input protection, zeroing and full scale adjust provisions, and input resistor balancing for minimum offset voltage.

Fig. 20-3

NANOAMPERE SENSING CIRCUIT WITH 100 MEGOHM INPUT IMPEDANCE



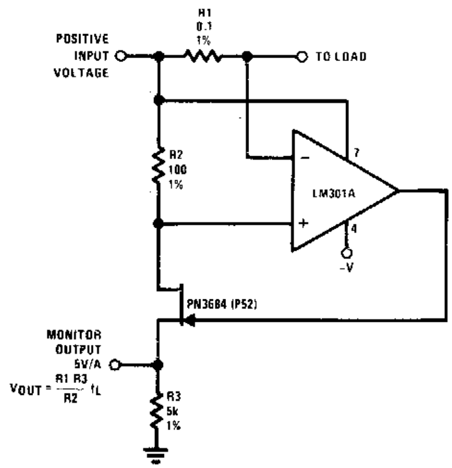
NANOAMPERE SENSING CIRCUIT WITH
100 MEGOHM INPUT IMPEDANCE

Fig. 20-4

Circuit Notes

The circuit may be used as a sensitive current detector or as a voltage detector having high input impedance. R1 is set so that the voltage at point (A) is ½ to ¾ volts below the level that fires the 2N494C. A small input current (Iin) of only 40 nanoamperes will charge C2 and raise the voltage at the emitter to the firing level. When the 2N494C fires, both capacitors, C1 and C2, are discharged through the 27 ohm resistor, which generates a positive pulse with sufficient amplitude to trigger a controlled rectifier (SCR), or other pulse sensitive circuitry.

CURRENT MONITOR



Circuit Notes

R1 senses current flow of a power supply. The JFET is used as a buffer because $I_D = I_S$; therefore the output monitor voltage accurately reflects the power supply current flow.

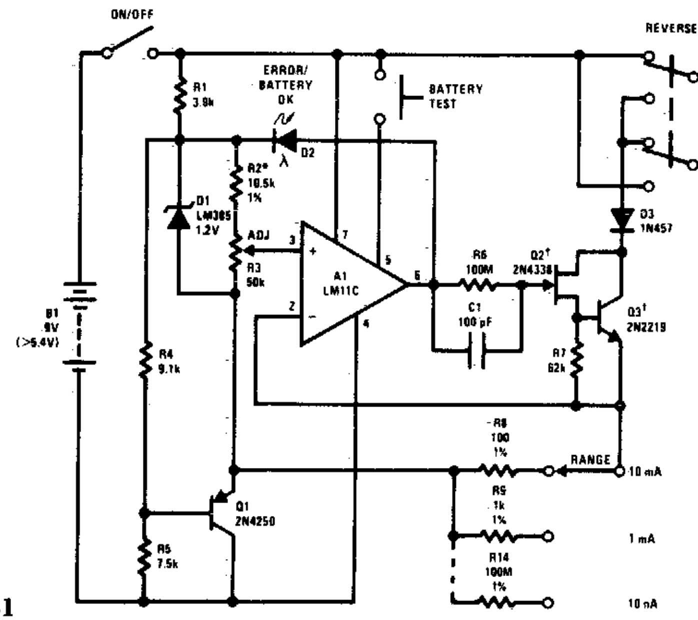
Fig. 20-5

21

Current Sources and Sinks

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

CURRENT SOURCE



* calibrate range

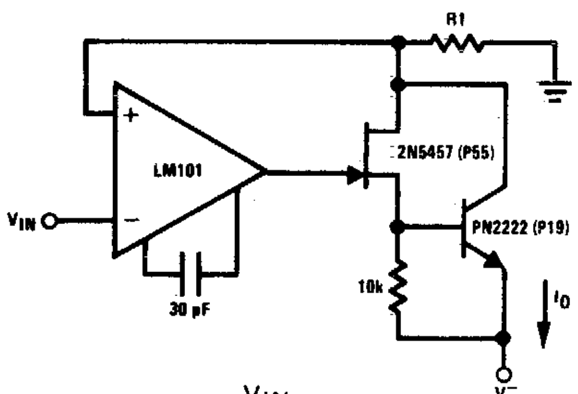
†select for $I_{CBO} \leq 100$ pA

Fig. 21-1

Circuit Notes

This precision current source has 10 μA to 10 mA ranges with output compliance of 30V to -5V. Output current is fully adjustable on each range with a calibrated, ten-turn potentiometer. Error light indicates saturation.

PRECISION CURRENT SOURCE



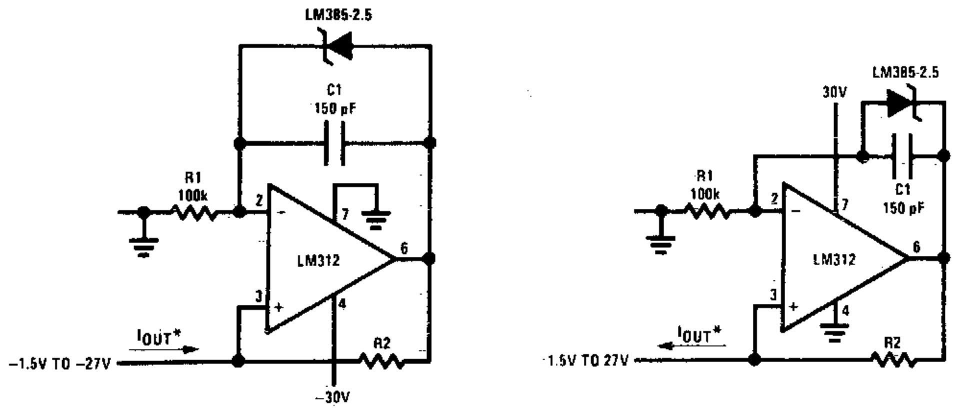
$$I_O = \frac{V_{IN}}{R1} \quad V_{IN} \leq 0V$$

Circuit Notes

The 2N5457 and PN2222 bipolar serve as voltage isolation devices between the output and the current sensing resistor, R1. The LM101 provides a large amount of loop gain to assure that the circuit acts as a current source. For small values of current (<1 mA), the PN2222 and 10K resistor may be eliminated with the output appearing at the source of the 2N5457.

Fig. 21-2

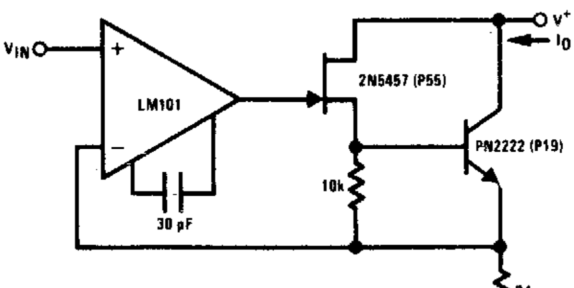
PRECISION 1 μA to 1 mA CURRENT SOURCES



$$^{*}I_{OUT} = \frac{2.5V}{R2}$$

Fig. 21-3

PRECISION CURRENT SINK



$$I_O = \frac{V_{IN}}{R1}$$

$$V_{IN} > 0V$$

Circuit Notes

The 2N5457 JFET and PN2222 bipolar have inherently high output impedance. Using R1 as a current sensing resistor to provide feedback to the LM101 op amp provides a large amount of loop gain for negative feedback to enhance the true current sink nature of this circuit. For small current values, the 10 K resistor and PN2222 may be eliminated if the source of the JFET is connected to R1.

Fig. 21-4

22

Dc/Dc and Dc/Ac Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Dc-to-Dc/Ac Inverter
Dc-to-Dc SMPS Using NE5561 Variable 18 V to 30 V Out at 0.2 A
Mini Power Inverter as High Voltage, Low Current Source
Regulated Dc-to-Dc Converter
400 V, 60 W Push-Pull Dc/Dc Converter
Dc/Dc Regulating Converter
Flyback Converter

DC-TO-DC/AC INVERTER

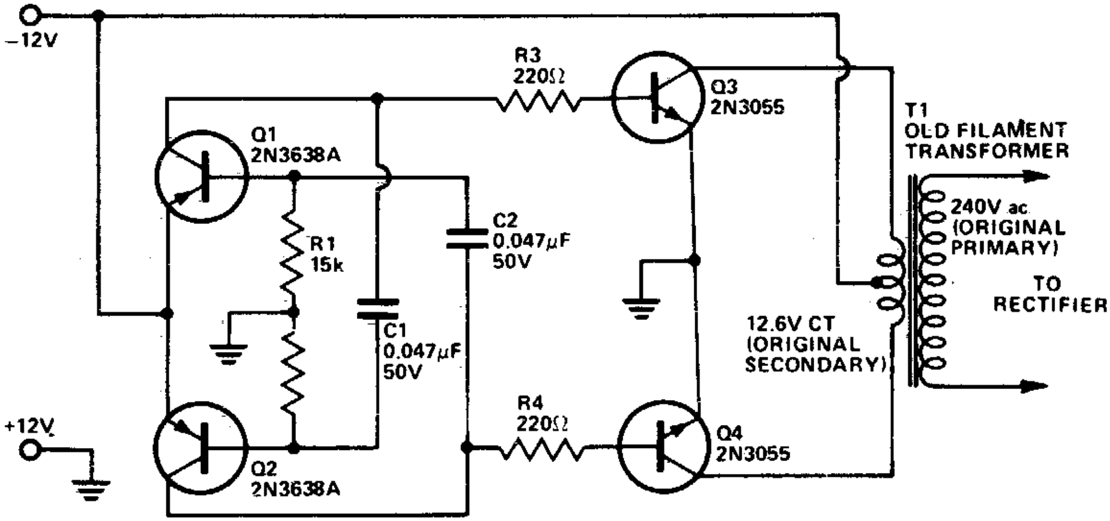
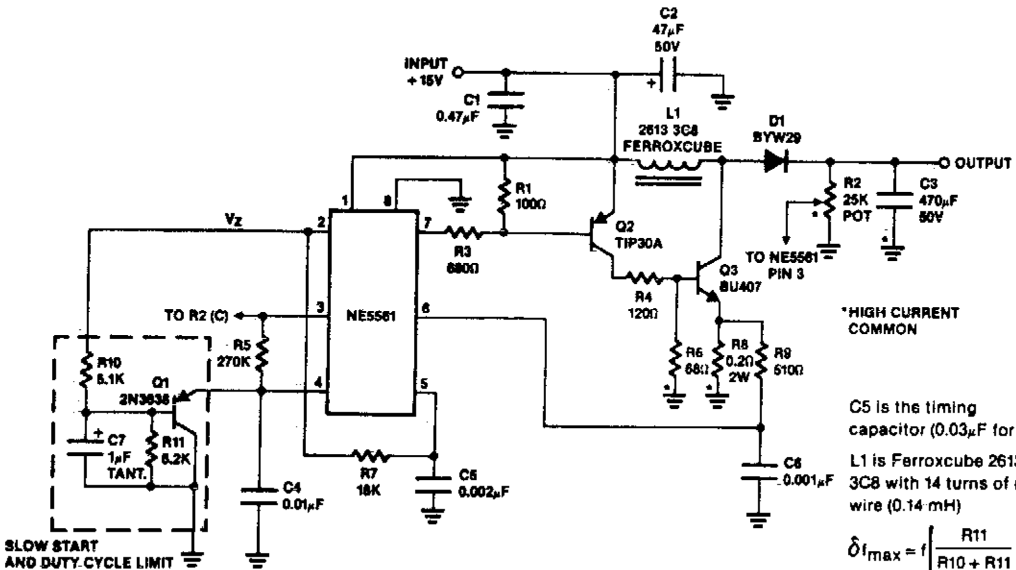


Fig. 22-1

Circuit Notes

This inverter uses no special components such as the torodial transformer used in many inverters. Cost is kept low with the use of cheap, readily available components. Essentially, it is a power amplifier driven by an astable multivibrator. The frequency is around 1200 Hz which most 50/60 Hz power transformers handle well without too much loss. Increasing the value of capacitors C1 and C2 will lower the frequency if any trouble is experienced. However, rectifier filtering capacitors required are considerably smaller at the higher operating frequency. The two 2N3055 transistor should be mounted on an adequately sized heatsink. The transformer should be rated according to the amount of output power required allowing for conversion efficiency of approximately 60%.

DC-TO-DC SMPS USING NE5561 VARIABLE 18 V to 30 V OUT AT 0.2 A



C5 is the timing capacitor (0.03µF for 22 kHz)

L1 is Ferroxcube 2613 3C8 with 14 turns of #18 enameled wire (0.14 mH)

$$\delta f_{max} = f\left[\frac{R11}{R10 + R11}(V_Z) + V_{BE}\right]$$

Fig. 22-2

MINI POWER INVERTER AS HIGH VOLTAGE, LOW CURRENT SOURCE

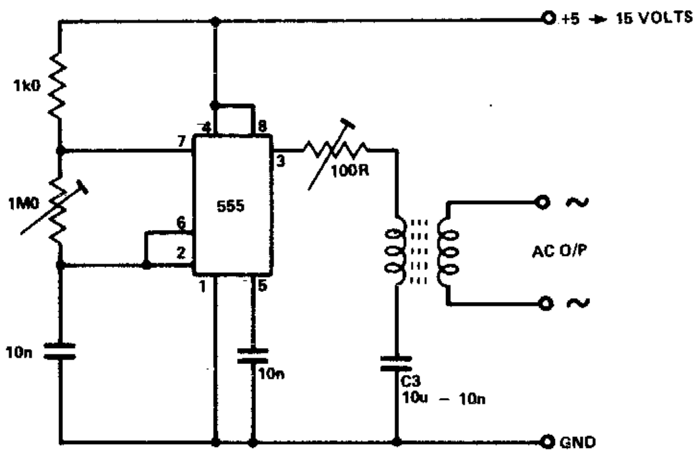


Fig. 22-3

Circuit Notes

The circuit is capable of providing power for portable Geiger counters, dosimeter chargers, high resistance meters, etc. The 555 timer IC is used in its multivibrator mode, the frequency adjusted to optimize the transformer characteristics. When the output of the IC is high, current flows through the limiting resistor, the primary coil to charge C3. When the output is low, the current is reversed. With a suitable choice of frequency and C3, a good symmetric output is sustained.

REGULATED DC-TO-DC CONVERTER

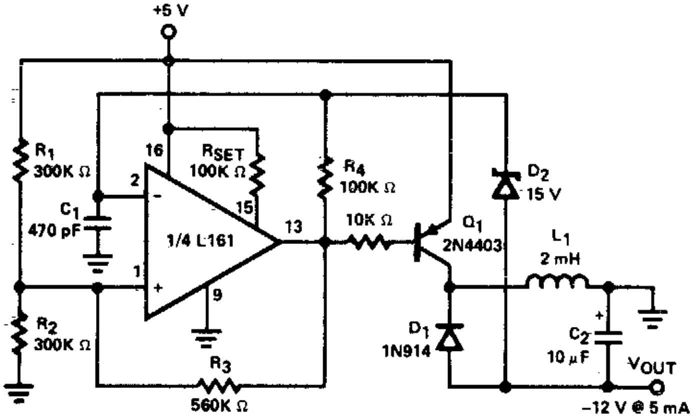


Fig. 22-4

Circuit Notes

Low power dc to dc converter obtained by adding a flyback circuit to a square wave oscillator. Operating frequency is 20 kHz to minimize the size of L1 and C2. Regulation is achieved by zener diode D2. Maximum current available before the converter drops out of regulation is 5.5 mA.

400 V, 60 W PUSH-PULL DC/DC CONVERTER

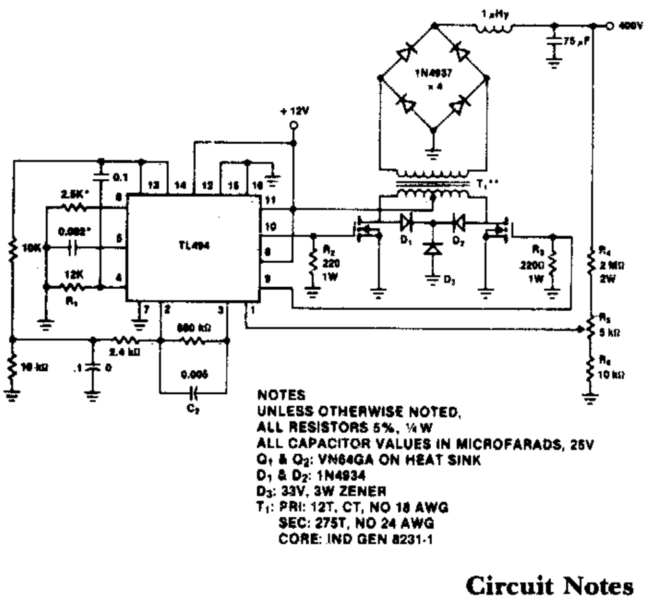


Fig. 22-5

Circuit Notes

The TL494 switching regulator governs the operating frequency and regulates output voltage. Switching frequency approximately 100 kHz for the values shown. Output regulation is typically 1.25% from no-load to full 60 W.

DC/DC REGULATING CONVERTER

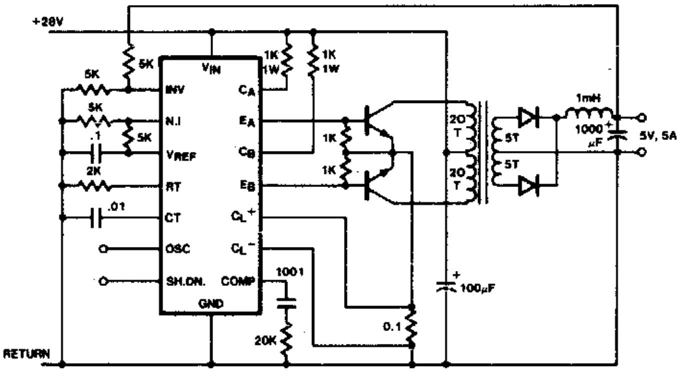


Fig. 22-6

Circuit Notes

Push-pull outputs are used in this transformer-coupled dc-dc regulating converter. Note that the oscillator must be set at twice the desired output frequency as the SG1524's internal flip-flop divides the frequency by 2 as it switches the PWM signal from one output to the other. Current limiting is done here in the primary so that the pulse width will be reduced should transformer saturation occur.

FLYBACK CONVERTER

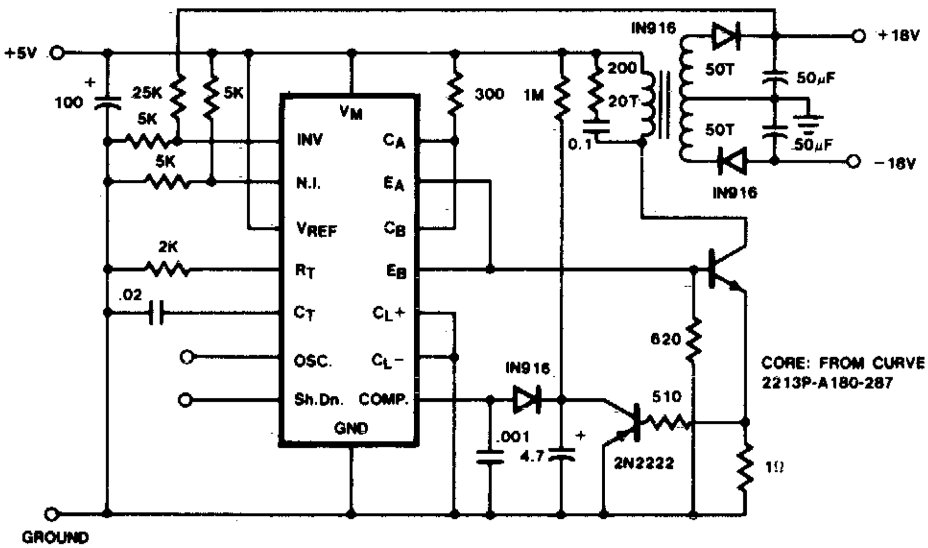


Fig. 22-7

Circuit Notes

A low-current flyback converter is used here to generate ±15 volts at 20 mA from a +5 volt regulated line. The reference generator in the SG1524 is unused with the input voltage providing the reference. Current limiting in a flyback converter is difficult and is accomplished here by sensing current in the primary line and resetting a soft-start circuit.

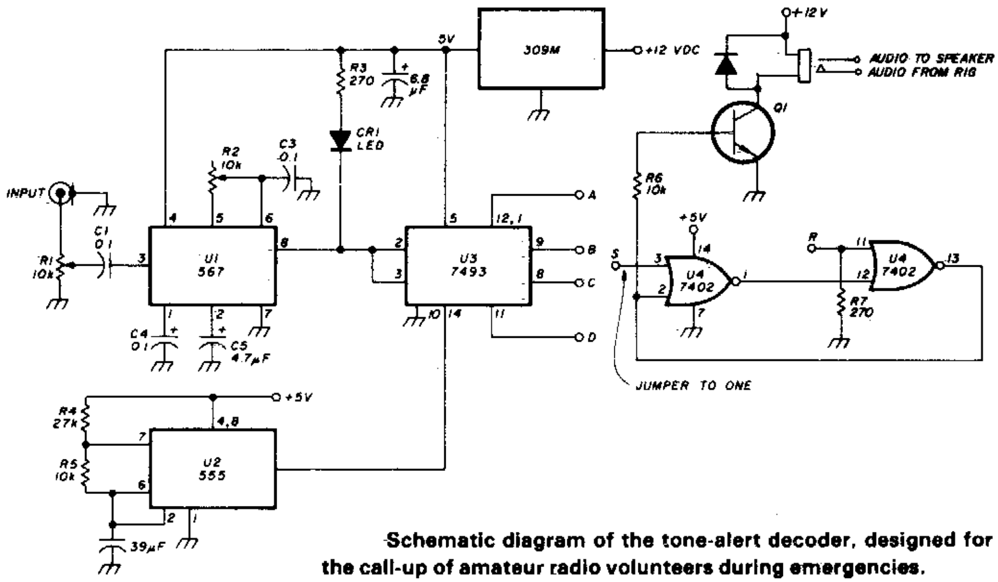
23

Decoders

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Tone Alert Decoder
Tone Decoder with Relay Output
SCA Decoder
10.8 MHz FSK Decoder
24% Bandwidth Tone Decoder
Dual-Tone Decoder

TONE-ALERT DECODER



Schematic diagram of the tone-alert decoder, designed for the call-up of amateur radio volunteers during emergencies.

Fig. 23-1

Circuit Notes

PLL (U1) is set with R2 to desired tone frequency. LED lights to indicate lock-up of PLL. Reduce signal level (R1) and readjust R2 to assure lock-up. Delay is selected from counter U3 output. Circuits latches (turns on Q1 to allow audio to speaker) when proper frequency/duration signal is received. To reset latch, a positive voltage must be applied briefly to the R input of U4.

TONE DECODER WITH RELAY OUTPUT

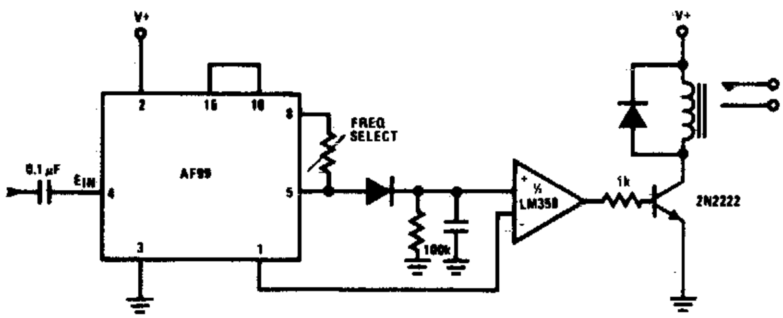


Fig. 23-2

SCA (BACKGROUND MUSIC) DECODER

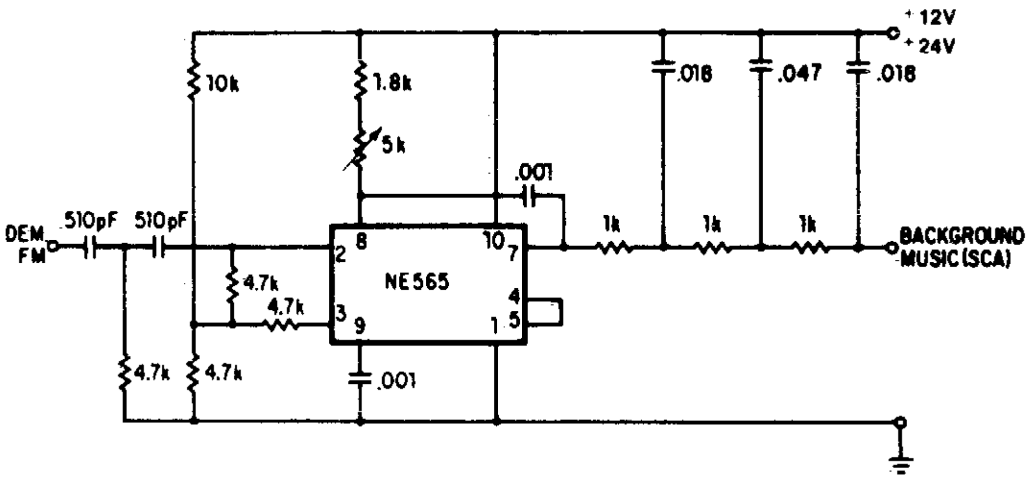


Fig. 23-3

Circuit Notes

A resistive voltage divider is used to establish a bias voltage for the input (pins 2 and 3). The demodulated (multiplex) FM signal is fed to the input through a two-stage high-pass filter, both to effect capacitive coupling and to attenuate the strong signal of the regular channel. A total signal amplitude, between 80 mV and 300 mV, is required at the input. Its source should have an impedance of less than 10,000 ohms. The Phase Locked Loop is tuned to 67 kHz with a 5000 ohm potentiometer; only approximate tuning is required, since the loop will seek the signal. The demodulated output (pin 7) passes through a three-stage low-pass filter to provide de-emphasis and attenuate the high-frequency noise which often accompanies SCA transmission. The demodulated output signal is in the rder of 50m V and the frequency response extends to 7 kHz.

10.8 MHz FSK DECODER

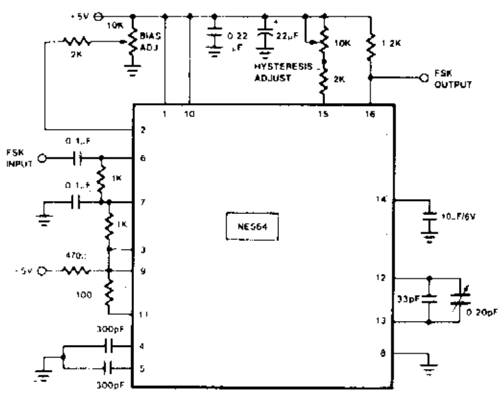


Fig. 23-4

24% BANDWIDTH TONE DECODER

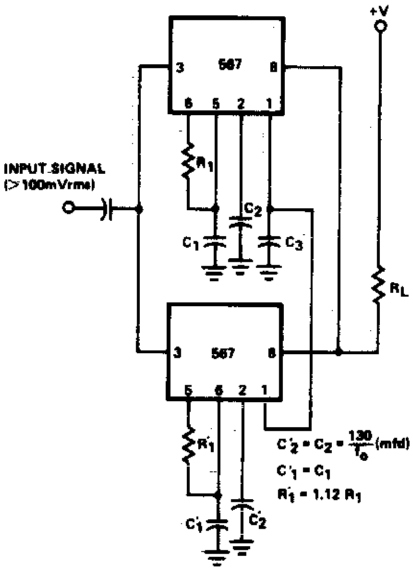


Fig. 23-5

DUAL-TONE DECODER

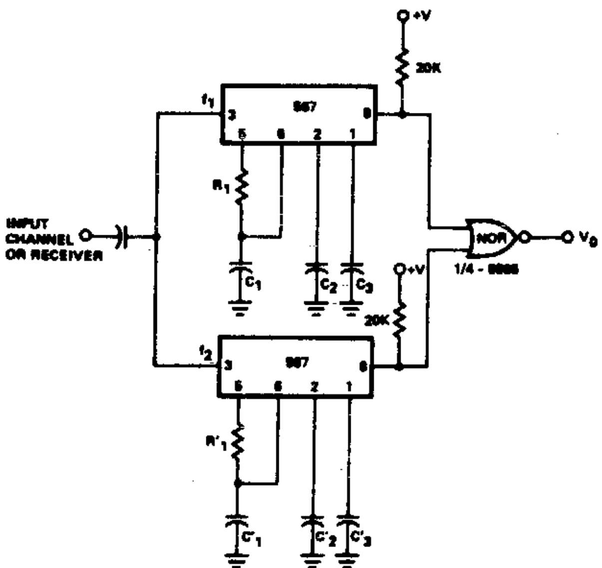


Fig. 23-6

1. Resistor and capacitor values chosen for desired frequencies and bandwidth.
2. If C_3 is made large so as to delay turn-on of the top 567, decoding of sequential (f_1 f_2) tones is possible.

24

Delays

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Long Time Delay
Time Delay Generator
Door Chimes Delay
Time Delay Generator
Long Delay Timer Using PUT
Ultra-Precise Long Time Delay Relay
Long Duration Time Delay
Simple Time Delay Using Two SCRs

LONG TIME DELAY

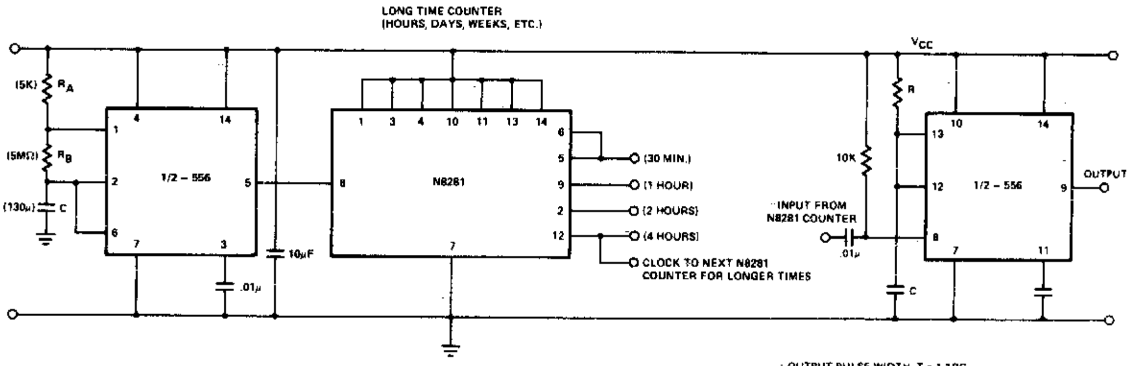


Fig. 24-1

Circuit Notes

In the 556 timer, the timing is a function of the charging rate of the external capacitor. For long time delays, expensive capacitors with extremely low leakage are required. The practicality of the components involved limits the time between pulses to something in the neighborhood of 10 minutes. To achieve longer time periods, both halves of a dual timer may be connected in tandem with a "Divide-by" network in between the first timer section operates in an oscillatory mode with a period of 1/fo. This signal is then applied to a "Divide-by-N" network to give an output with the period of N/fo. This can then be used to trigger the second half of the 556. The total time delay is now a function of N and fo.

TIME DELAY GENERATOR

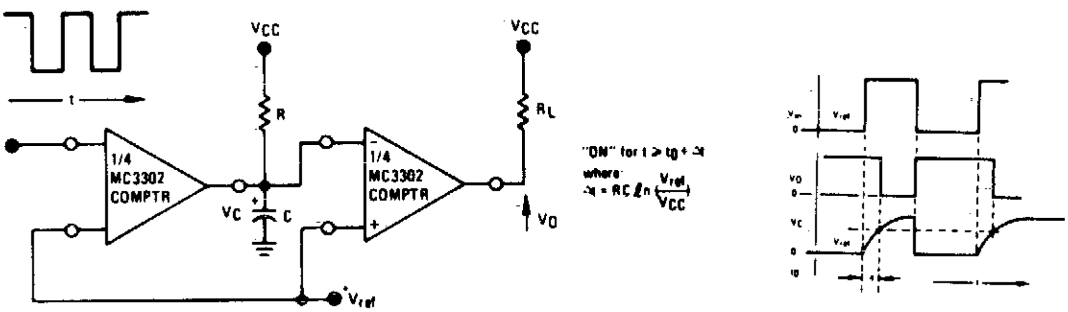


Fig. 24-2

DOOR CHIMES DELAY

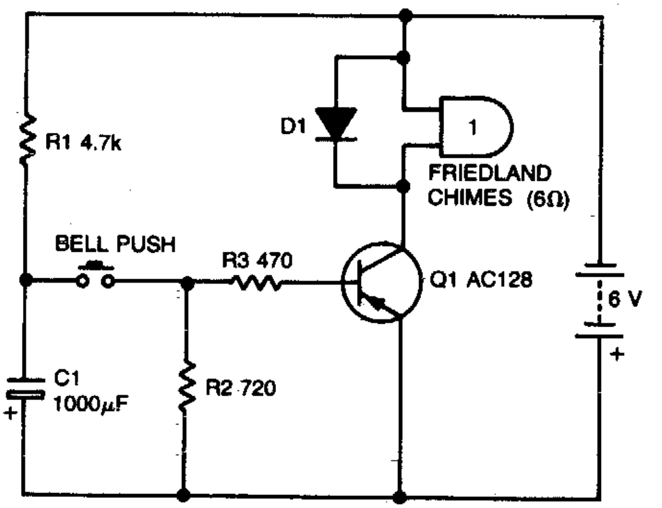


Fig. 24-3

Circuit Notes

With values shown, this simple circuit will permit one operation every 10 seconds or so. Capacitor C1 charges through R1 when the button is released. Making R1 larger will increase the delay.

TIME DELAY GENERATOR

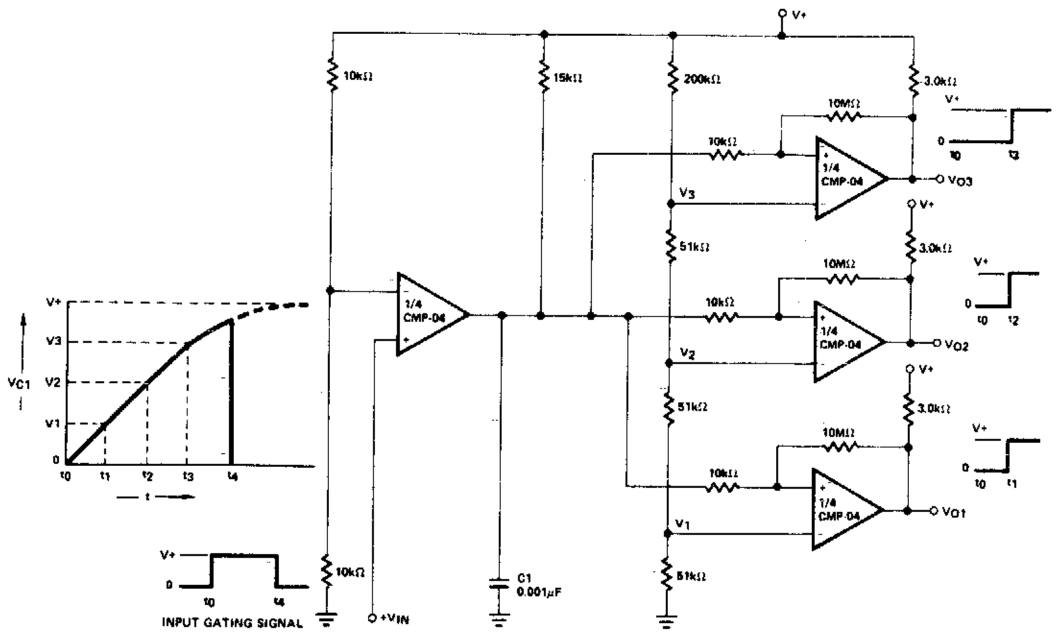


Fig. 24-4

LONG DELAY TIMER USING PUT

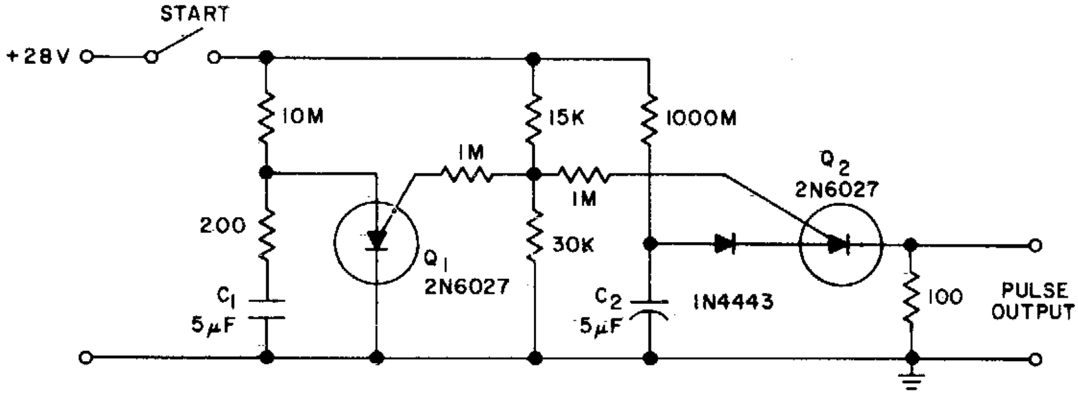


Fig. 24-5

Circuit Notes

The PUT is used as both a timing element and sampling oscillator. A low leakage film capacitor is required for C2 due to the low current supplied to it.

ULTRA-PRECISE LONG TIME DELAY RELAY

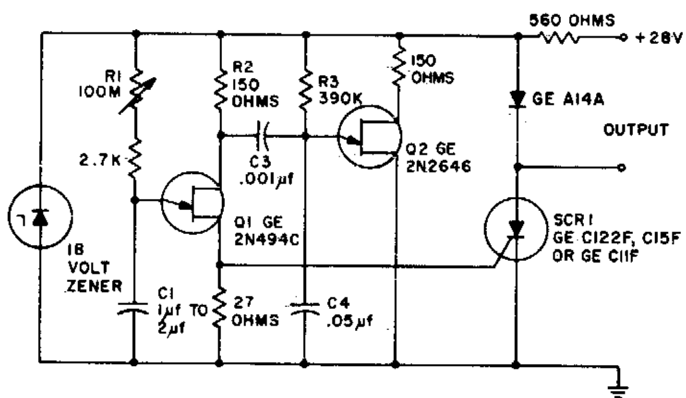


Fig. 24-6

Circuit Notes

Predictable time delays from as low as 0.3 milliseconds to over 3 minutes are obtainable without resorting to a large value electrolytic-type timing capacitor. Instead, a stable low leakage paper or mylar capacitor is used and the peak point current of the timing UJT (Q1) is effectively reduced, so that a large value emitter resistor (R1) may be substituted.

LONG DURATION TIME DELAY

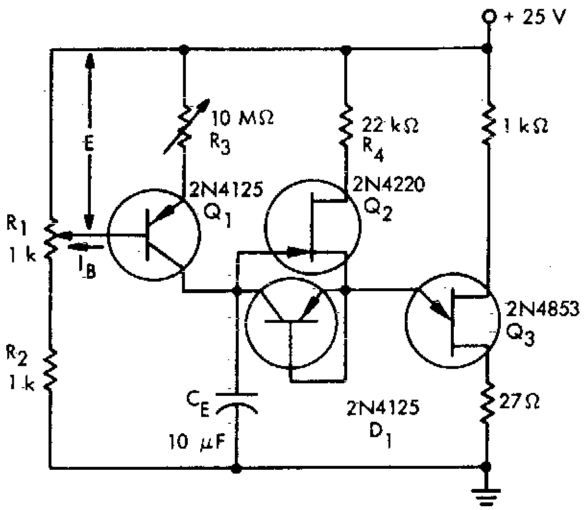


Fig. 24-7

SIMPLE TIME DELAY

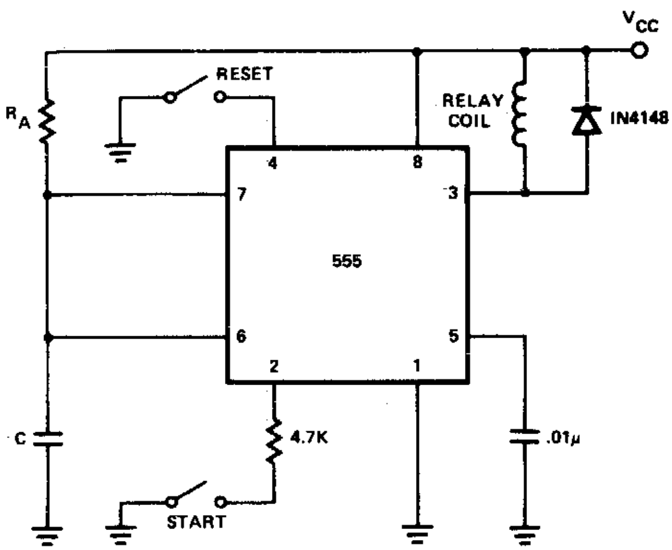


Fig. 24-8

25

Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AIR-MOTION DETECTOR

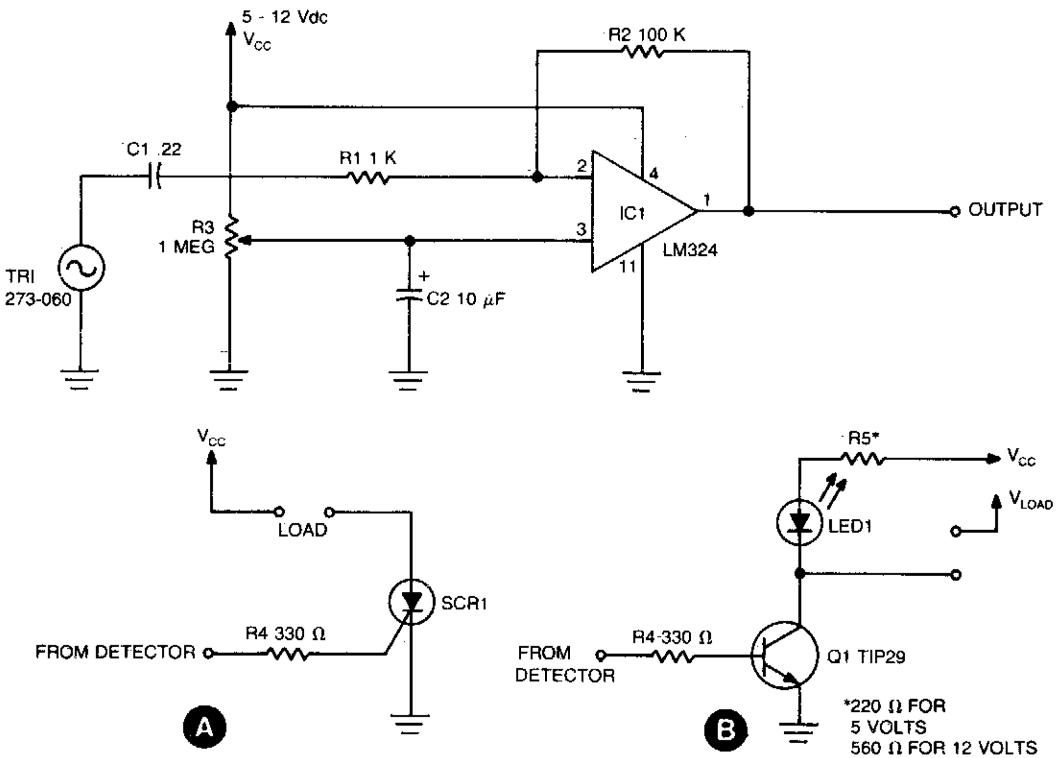


Fig. 25-1

Circuit Notes

Sensing circuit detects either steady or fluctuating air flows. The heart of the circuit is a Radio Shack piezo buzzer (P/N 273-060) and an LM324 quad op amp. (Red wire from the piezo element connects to capacitor C1, and the black wire to ground.) When a current of air hits the piezo element, a small signal is generated and is fed through C1 and R1 to the inverting input (pin 2) of one section of the LM324. That causes the output (pin 1) to go high. Resistor R3 adjusts sensitivity. The circuit can be made sensitive enough to detect the wave of a hand or the sensitivity can be set so low that blowing on the element hard will produce no output. Resistor R2 is used to adjust the level of the output voltage at pin 1. The detector circuit can be used in various control applications. For example, an SCR can be used to control 117-volt AC loads as shown in A. Also, an NPN transistor, such as a TIP29, can be used to control loads as shown in B.

PRODUCT DETECTOR

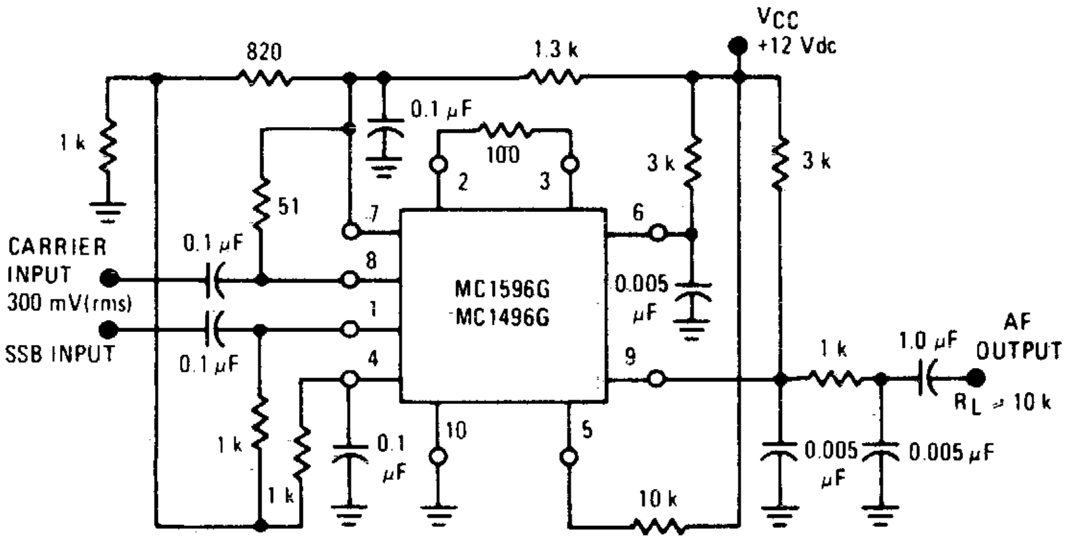


Fig. 25-2

Circuit Notes

The MC1596/MC1496 makes an excellent SSB product detector. This product detector has a sensitivity of 3.0 microvolts and a dynamic range of 90 dB when operating at an intermediate frequency of 9 MHz. The detector is broadband for the entire high frequency range. For operation at very low intermediate frequencies down to 50 kHz the 0.1 μF capacitors on pins 7 and 8 should be increased to 1.0 μF. Also, the output filter at pin 9 can be tailored to a specific intermediate frequency and audio amplifier input impedance. The emitter resistance between pins 2 and 3 may be increased or decreased to adjust circuit gain, sensitivity, and dynamic range. This circuit may also be used as an AM detector by introducing carrier signal at the carrier input and an AM signal at the SSB input. The carrier signal may be derived from the intermediate frequency signal or generated locally. The carrier signal may be introduced with or without modulation, provided its level is sufficiently high to saturate the upper quad differential amplifier. If the carrier signal is modulated, a 300 mV (rms) input level is recommended.

LOW VOLTAGE DETECTOR

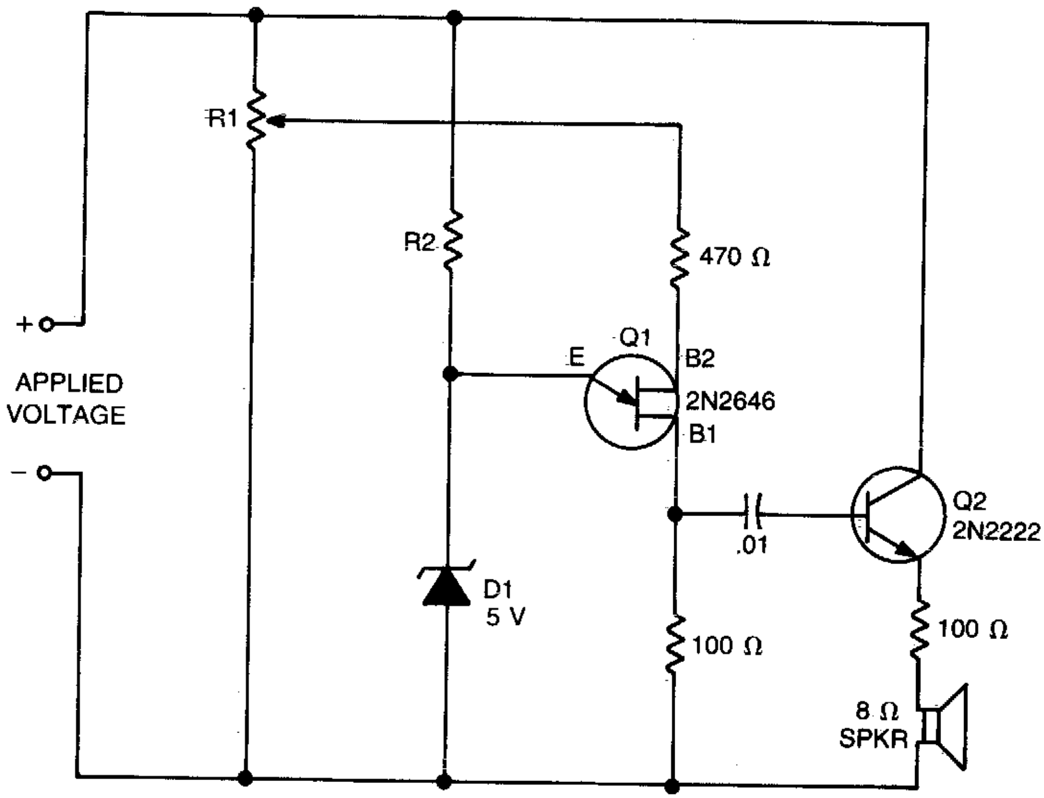


Fig. 25-3

Circuit Notes

The values of R1, R2, and D1 are selected for the voltage applied. Using a 12-volt battery, R1 = 10 K, R2 = 5.6 K and D1 is a 5-volt zener diode, or a string of forward-biased silicon rectifiers equaling about 5 volts. Transistor Q1 is a general-purpose UJT (Unijunction Transistor), and Q2 is any small-signal or switching NPN transistor. When detector is connected across the battery terminals, it draws little current and does not interfere with other devices powered by the battery. If voltage drops below the trip voltage selected with the R1 setting, the speaker beeps a warning. The frequency of the beeps is determined by the amount of undervoltage. If other voltages are being monitored, select R1 so that it draws only 1 mA or 2 mA. Zener diode D1 is about one-half of the desired trip voltage, and R2 is selected to bias it about 1 mA.

POSITIVE PEAK DETECTOR

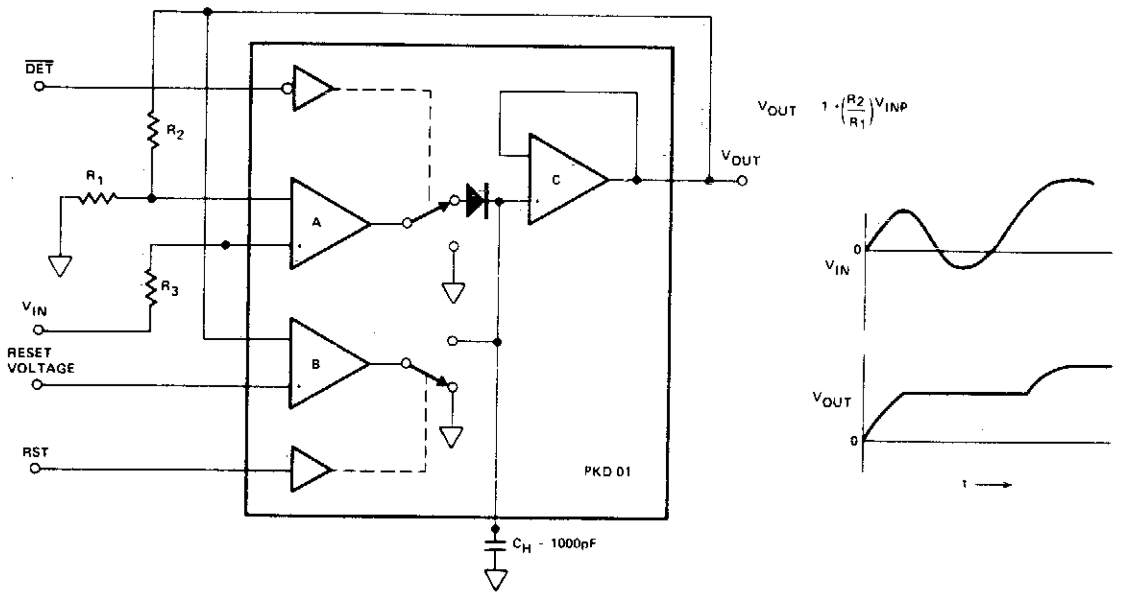


Fig. 25-4

NEGATIVE PEAK DETECTOR

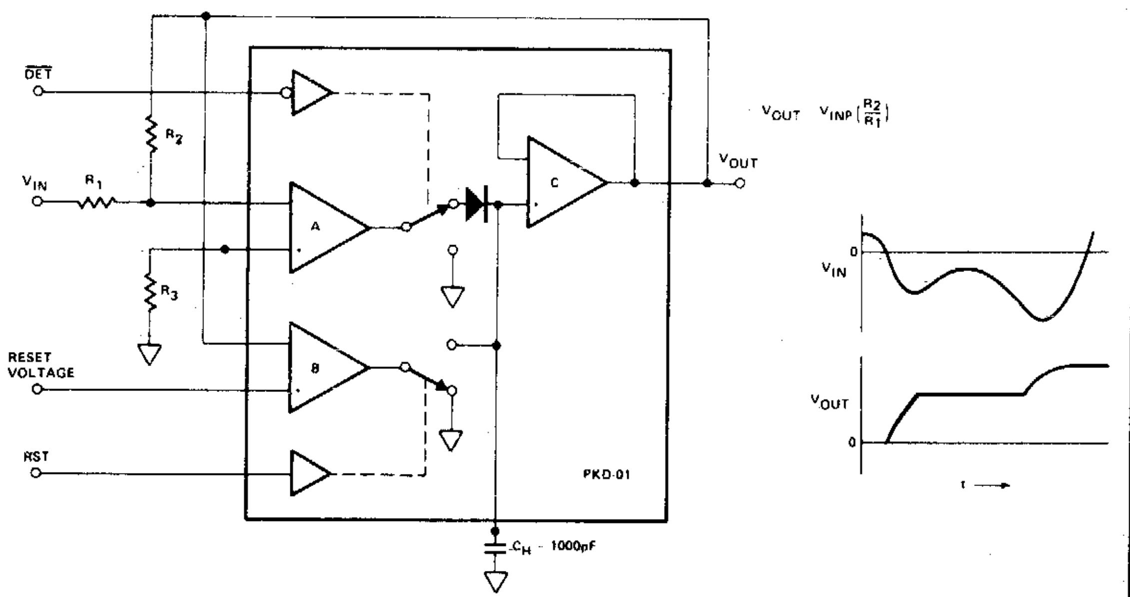
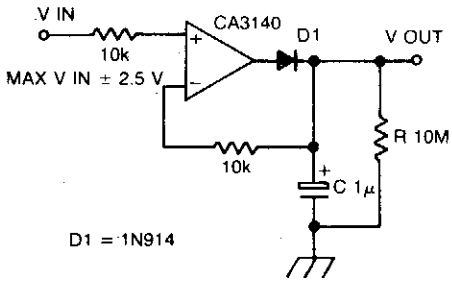


Fig. 25-5

PRECISION PEAK VOLTAGE DETECTOR WITH A LONG MEMORY TIME



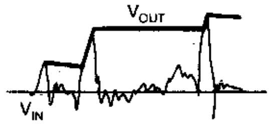


Fig. 25-6

Circuit Notes

The circuit has negative feedback only for positive signals. The inverting input can only get some feedback when diode D1 is forward biased and only occurs when the input is positive. With a positive input signal, the output of the op amp rises until the inverting input signal reaches the same potential. In so doing, the capacitor C is also charged to this potential. When the input goes negative, the diode D1 becomes reverse biased, the voltage on the capacitor remains, being slowly discharged by the op amp input bias current of 10 pico amps. Thus the discharge of the capacitor is dominantly controlled by the resistor R, giving a time constant of 10 seconds. Thus, the circuit detects the most positive peak voltage and remembers it.

EDGE DETECTOR

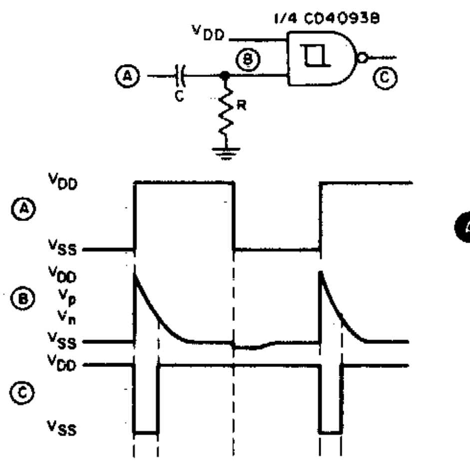
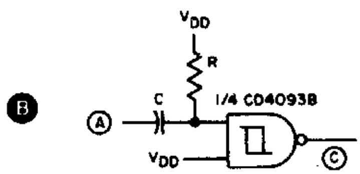


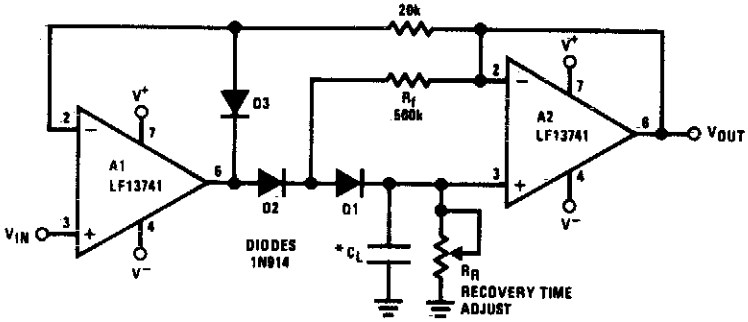
Fig. 25-7

Circuit Notes

This circuit provides a short negative-going output pulse for every positive-going edge at the input. The input waveform is coupled to the input by capacitor C; the pulse length depends, as before, on R and C. If a negative going edge detector is required, the circuit in B should be used.



ULTRA-LOW DRIFT PEAK DETECTOR

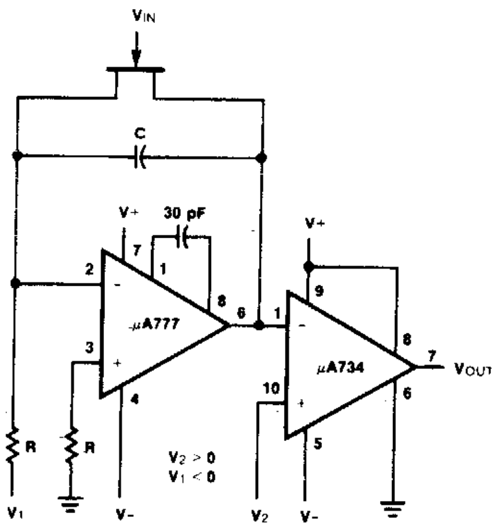


- By adding D1 and R_f, $V_{D1} = 0$ during hold mode. Leakage of D2 provided by feedback path through R_f.
- Leakage of circuit is I_B plus leakage of C_h.
- D3 clamps V_{OUT} A1 to $V_{IN} - V_{D3}$ to improve speed and to limit the reverse bias of D2.
- Maximum input frequency should be $<< 1/2\pi R_f C_{D2}$, where C_{D2} is the shunt capacitance of D2.

*Low leakage capacitor

Fig. 25-8

PULSE WIDTH DISCRIMINATOR

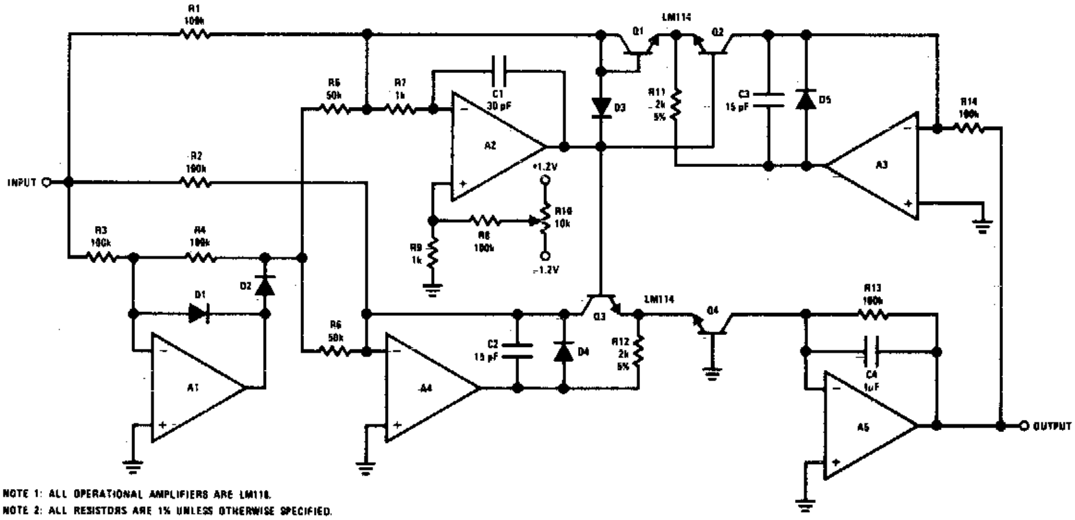


V_{OUT} Pulse Appears

Whenever $T > \frac{-RCV_2}{V_1}$

Fig. 25-9

TRUE RMS DETECTOR



NOTE 1: ALL OPERATIONAL AMPLIFIERS ARE LM118.
NOTE 2: ALL RESISTORS ARE 1% UNLESS OTHERWISE SPECIFIED.
NOTE 3: ALL DIODES ARE 1N914.
NOTE 4: SUPPLY VOLTAGE ±15V.

Fig. 25-10

Circuit Notes

The circuit will provide a dc output equal to the rms value of the input. Accuracy is typically 2% for a 20 V_{PP} input signal from 50 Hz to 100 kHz, although it's usable to about 500 kHz. The lower frequency is limited by the size of the filter capacitor. Since the input is dc coupled, it can provide the true rms equivalent of a dc and ac signal.

FAST HALF-WAVE RECTIFIER

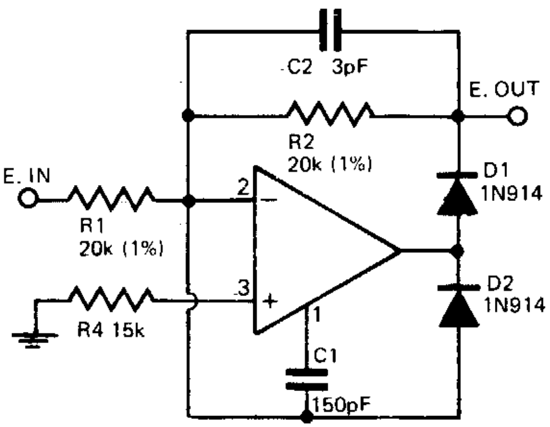


Fig. 25-11

Circuit Notes

Precision half wave rectifier using an operational amplifier will have a rectification accuracy of 1% from dc to 100 kHz.

TELEMETRY DEMODULATOR

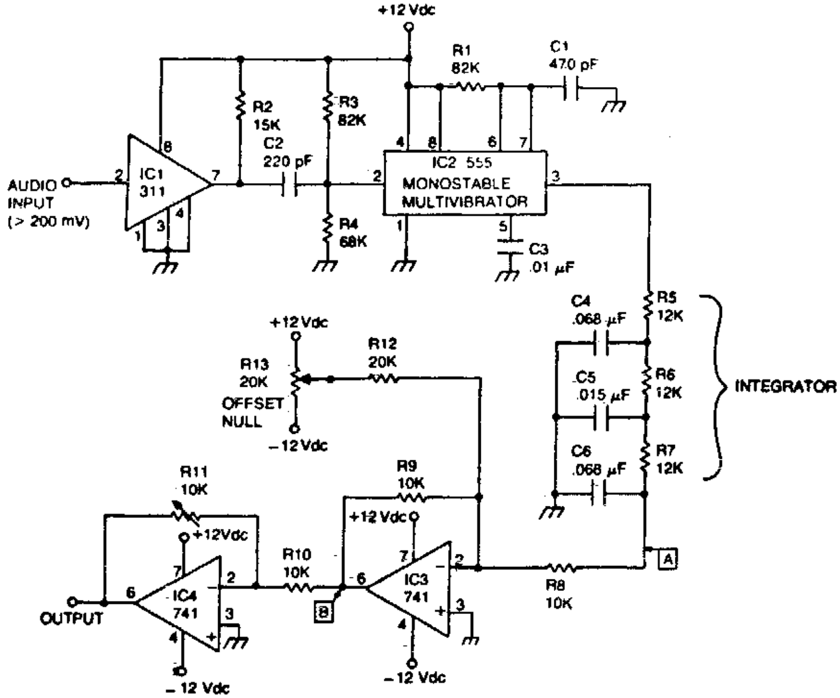


Fig. 25-12

Circuit Notes

The circuit recovers an FM audio signal that varies from less than 1 kHz to about 10 kHz.

FULL-WAVE RECTIFIER AND AVERAGING FILTER

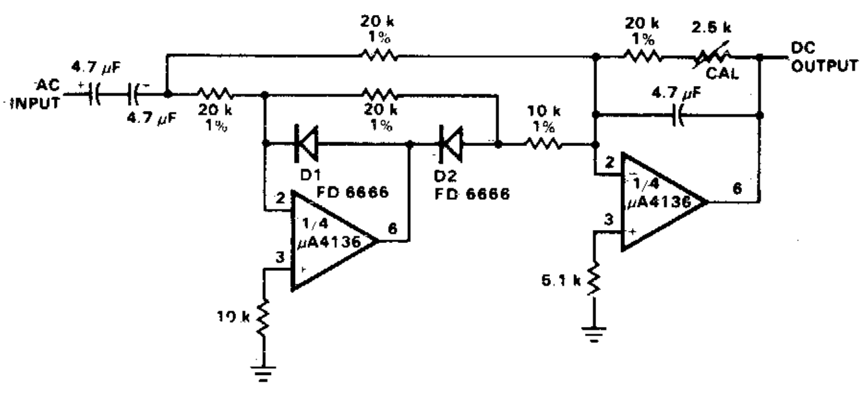
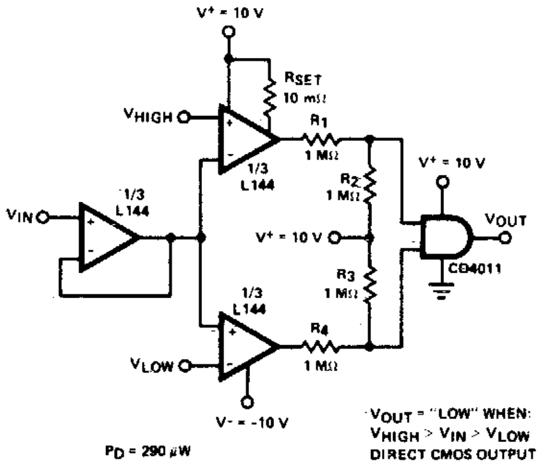


Fig. 25-13

DOUBLE-ENDED LIMIT DETECTOR



Circuit Notes

Detector uses three sections of an L144 and a CMOS NAND gate to make a very low power voltage monitor. The 1 MΩ resistors R1, R2, R3, and R4 translate the bipolar ±10 V swing of the op amps to a 0 to 10 V swing acceptable to the ground-referenced CMOS logic. The total power dissipation is 290 μW while in limit and 330 μW while out of limit.

Fig. 25-14

HALF-WAVE RECTIFIER

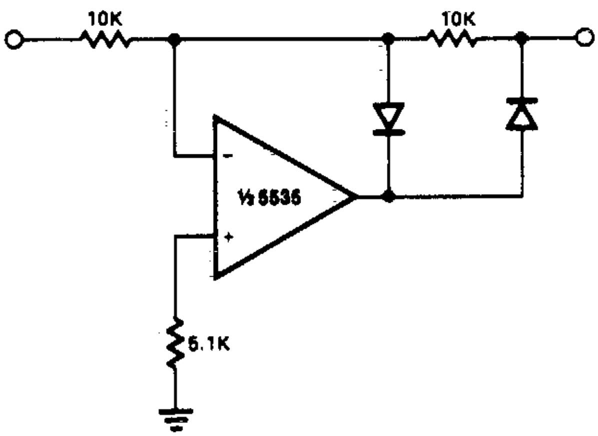


Fig. 25-15

All resistor values are in ohms.

Circuit Notes

This circuit provides for accurate half wave rectification of the incoming signal. For positive signals, the gain is 0; for negative signals, the gain is −1. By reversing both diodes, the polarity can be inverted. This circuit provides an accurate output, but the output impedance differs for the two input polarities and buffering may be needed. The output must slew through two diode drops when the input polarity reverses. The NE5535 device will work up to 10 kHz with less ttan 5% distortion.

TONE DETECTOR

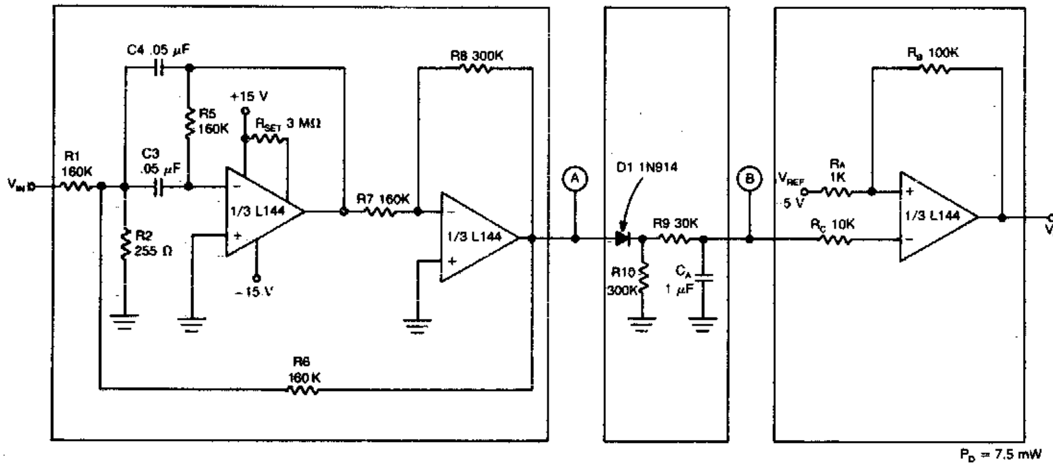
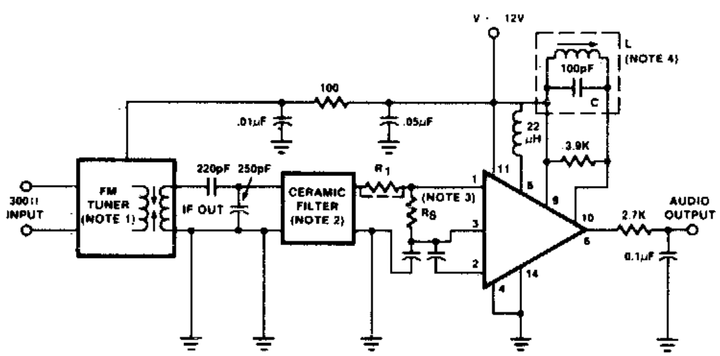


Fig. 25-16

Circuit Notes

The detector circuit is made up a two-amplifier multiple feedback bandpass filter followed by an ac-to-dc detector section and a Schmitt Trigger. The bandpass filter (with a Q of greater than 100) passes only 500 Hz inputs whch are in turn rectified by D1 and filtered by R9 and CA. This filtering action in combination with the trigger level of 5 V for the Schmitt device insures that at least 55 cycles of 500 Hz input must be present before the output will react to a tone input.

FM TUNER WITH A SINGLE-TUNED DETECTOR COIL



NOTES

All resistors values are typical and in ohms.

1. Weller 4SN3FIC or equivalent
2. Murate SFG 10.7mA or equivalent
3. R_S will affect stability depending on circuit layout. To increase stability R_S is decreased. Range of R_S is 330 to 50Ω, $R_1 + R_S \leq 330Ω$
4. L tunes with 100pF (C) at 10.7MHz Q_O unloaded ≅ 75 (G.I. EX27825 or equivalent)

Performance data at f_o = 98MHz, f_{MOD} = 400Hz, deviation = ± 74kHz:

−3dB limiting sensitivity	2μV (antenna level)
20dB quieting sensitivity	1μV (antenna level)
30dB quieting sensitivity	1.5μV (antenna level)

Fig. 25-17

MISSING PULSE DETECTOR

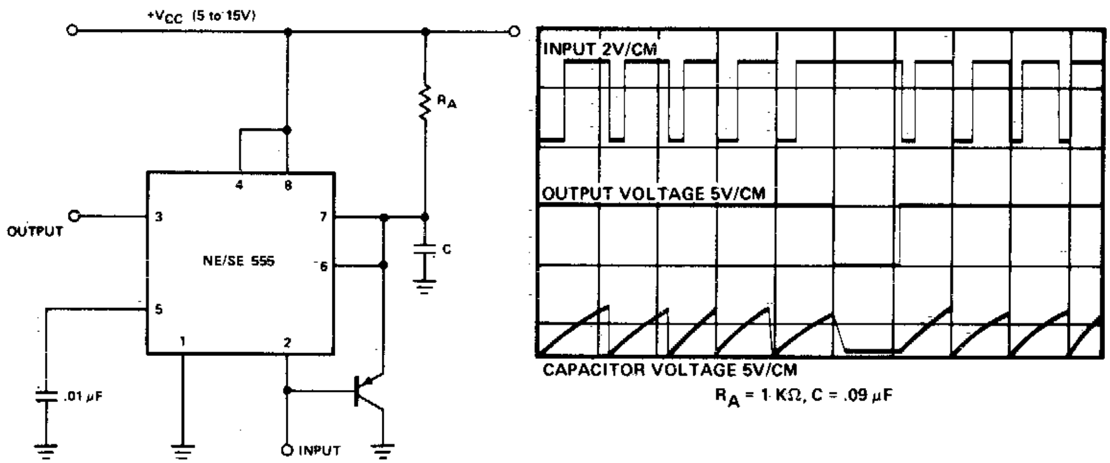


Fig. 25-18

Circuit Notes

The timing cycle is continuously reset by the input pulse train. A change in frequency, or a missing pulse, allows completion of the timing cycle which causes a change in the output level. For this application, the time delay should be set to be slightly longer than the normal time between pulses. The graph shows the actual waveforms seen in this mode of operation.

HIGH SPEED PEAK DETECTOR

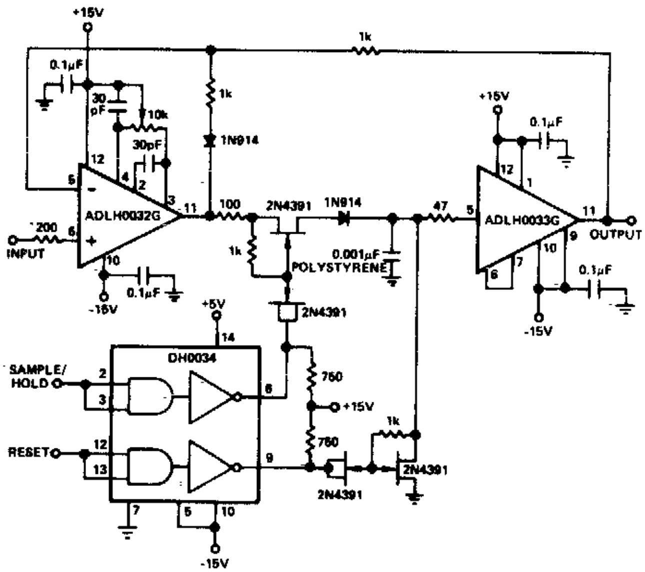


Fig. 25-19

DETECTOR FOR MAGNETIC TRANSDUCER

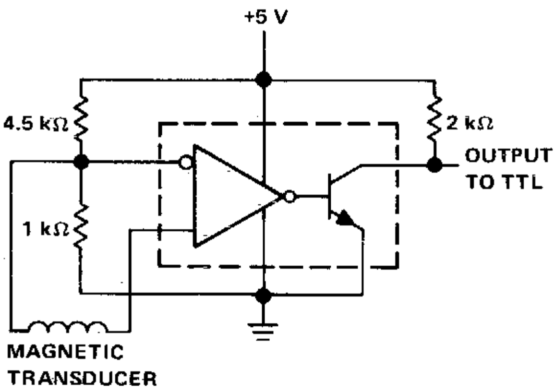


Fig. 25-20

FM DEMODULATOR AT 5 V

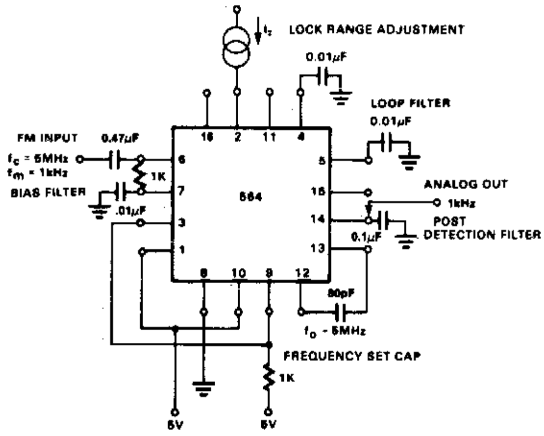


Fig. 25-22

DOUBLE-ENDED LIMIT DETECTOR

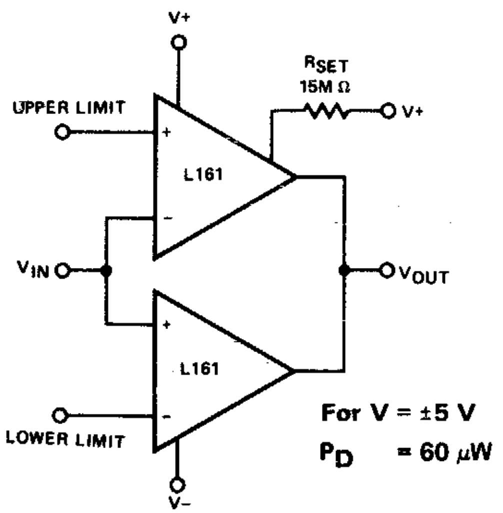


Fig. 25-21

FM DEMODULATOR AT 12 V

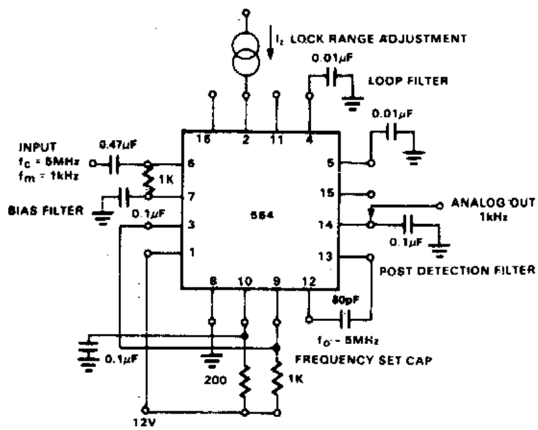


Fig. 25-23

PRECISION FULL WAVE RECTIFIER

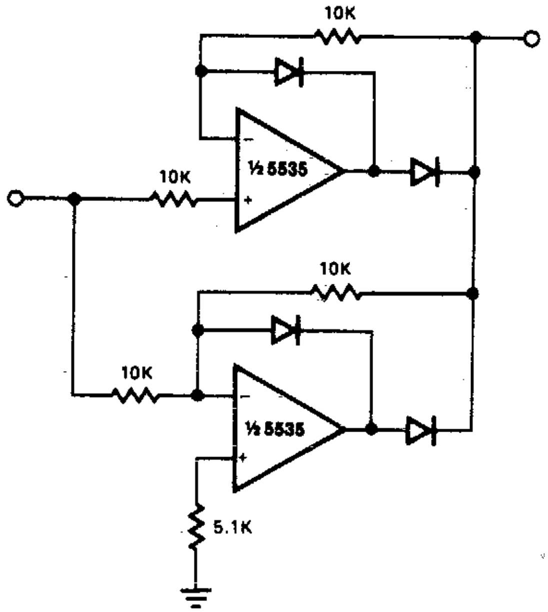


Fig. 25-24

Circuit Notes

The circuit provides accurate full wave rectification. The output impedance is low for both input polarities, and the errors are small at all signal levels. Note that the output will not sink heavy current, except a small amount through the 10 K resistors. Therefore, the load applied should be referenced to ground or a negative voltage. Reversal of all diode polarities will reverse the polarity of the output. Since the outputs of the amplifiers must slew through two diode drops when the input polarity changes, 741 type devices give 5% distortion at about 300 Hz.

NEGATIVE PEAK DETECTOR

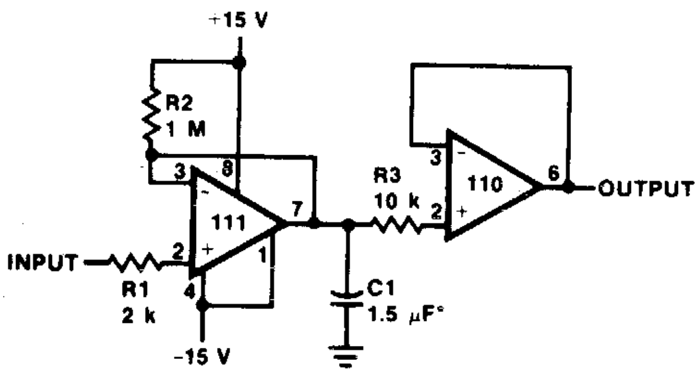


Fig. 25-25

*Solid tantalum

LEVEL DETECTOR WITH HYSTERESIS (POSITIVE FEEDBACK)

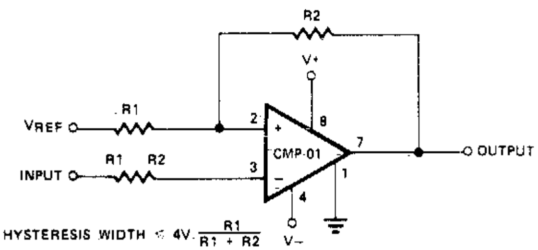
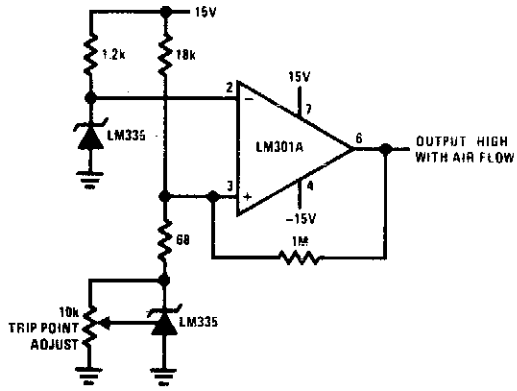


Fig. 25-26

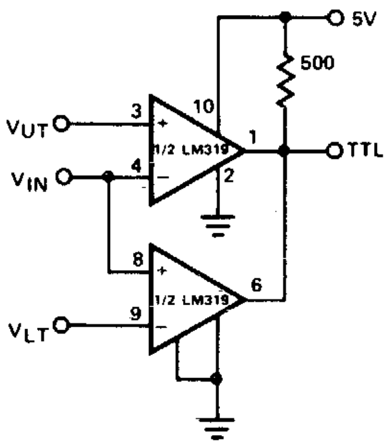
AIR FLOW DETECTOR



*Self heating is used to detect air flow

Fig. 25-28

WINDOW DETECTOR

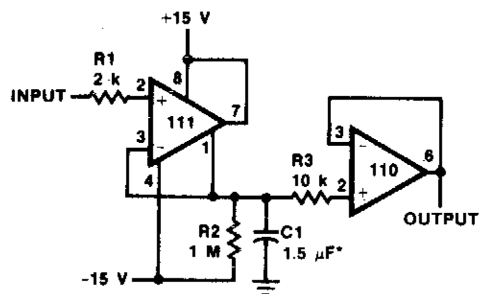


$V_{OUT} = 5V$ for $V_{LT} < V_{IN} < V_{UT}$

$V_{OUT} = 0$ for $V_{IN} < V_{LT}$ or $V_{IN} > V_{UT}$

Fig. 25-27

POSITIVE PEAK DETECTOR



*Solid tantalum

Fig. 25-29

26

Digital-to-Analog Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

14-BIT BINARY D/A CONVERTER (UNIPOLAR)

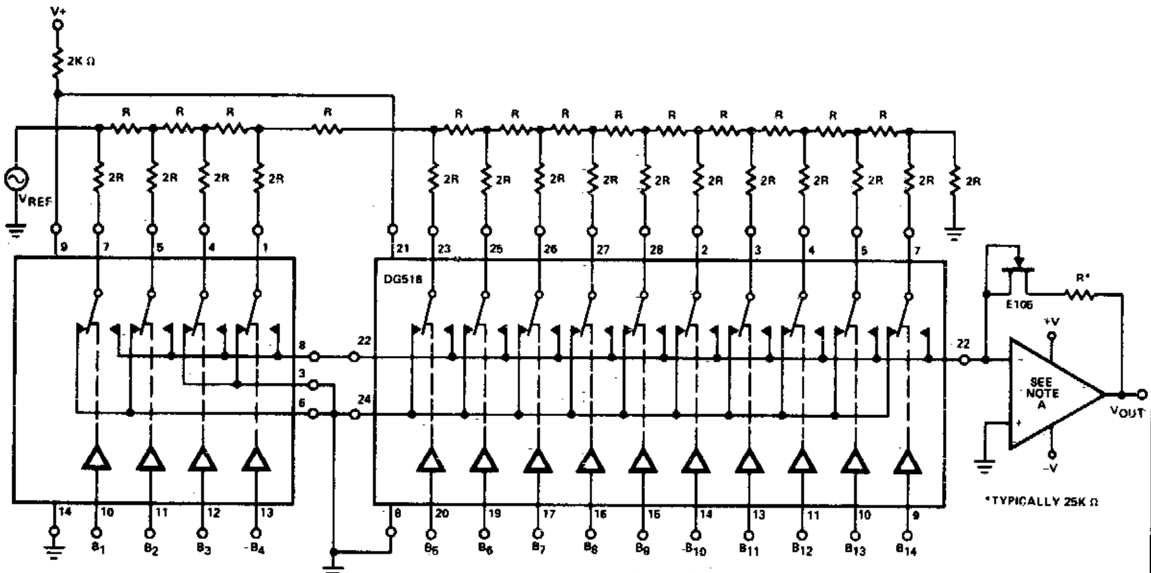


Figure 7

NOTE:

A. Op-Amp characteristics effect D/A accuracy and settling time. The following Op-Amps, listed in order of increasing speed, are suggested:

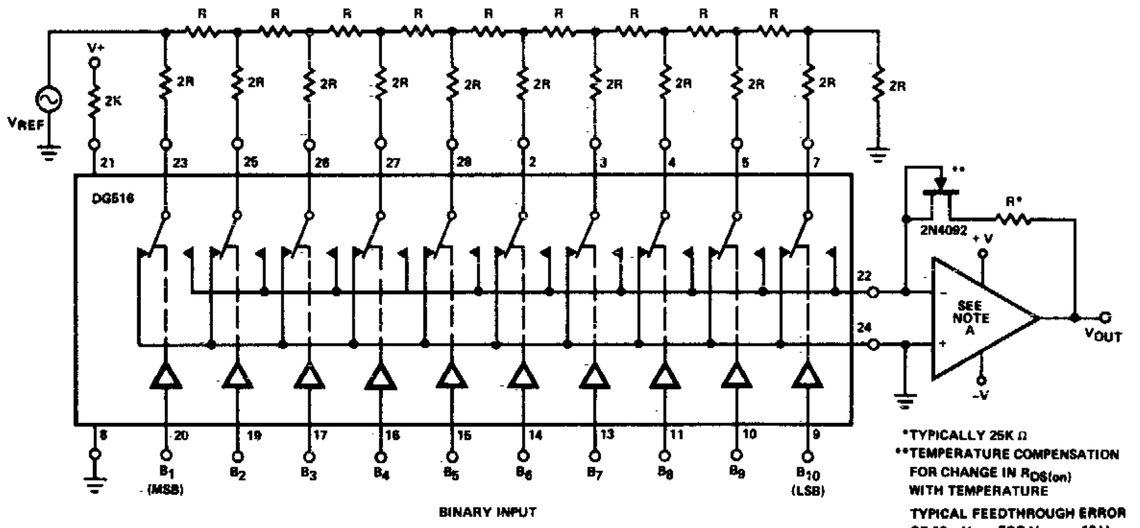
1. LM101A
2. LF156A
3. LM118

Unipolar Binary Operation

DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1 1 1 1 1	$-V_{REF}(1-2^{-14})$
1 0 0 0 0 0 0 0 0 0 0 0 0 1	$-V_{REF}(1/2+2^{-14})$
1 0 0 0 0 0 0 0 0 0 0 0 0 0	$-V_{REF}/2$
0 1 1 1 1 1 1 1 1 1 1 1 1 1	$-V_{REF}(1/2-2^{-14})$
0 0 0 0 0 0 0 0 0 0 0 0 0 1	$-V_{REF}(2^{-14})$
0 0 0 0 0 0 0 0 0 0 0 0 0 0	0

Fig. 26-1

10 BIT D/A CONVERTER



Unipolar Binary Operation

DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1	$-V_{REF}\ (1 - 2^{-10})$
1 0 0 0 0 0 0 0 0 1	$-V_{REF}\ (1/2 + 2^{-10})$
1 0 0 0 0 0 0 0 0 0	$-V_{REF}/2$
0 1 1 1 1 1 1 1 1 1	$-V_{REF}\ (1/2 - 2^{-10})$
0 0 0 0 0 0 0 0 0 1	$-V_{REF}\ (2^{-10})$
0 0 0 0 0 0 0 0 0 0	0

NOTE:

Op-Amp characteristics effect D/A accuracy and settling time. The following Op-Amps, listed in order of increasing speed, are suggested:

1. LM101A
2. LF156A
3. LM118

Fig. 26-2

FAST VOLTAGE OUTPUT D/A CONVERTER

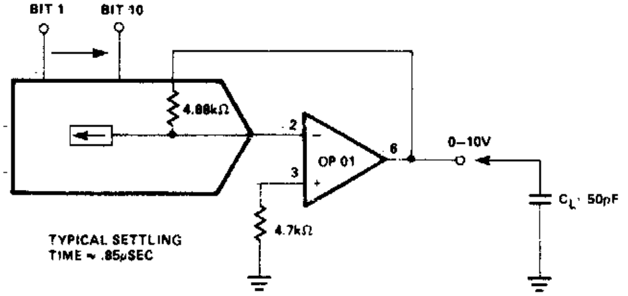


Fig. 26-3

RESISTOR TERMINATED DAC (0 TO −5 V OUTPUT)

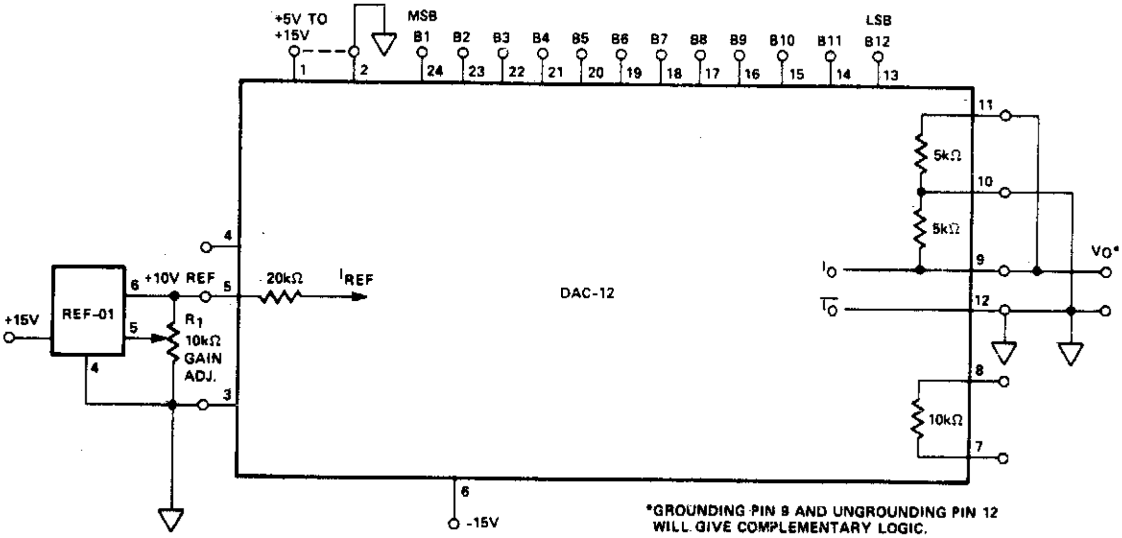


Fig. 26-4

THREE-DIGIT BCD D/A CONVERTER

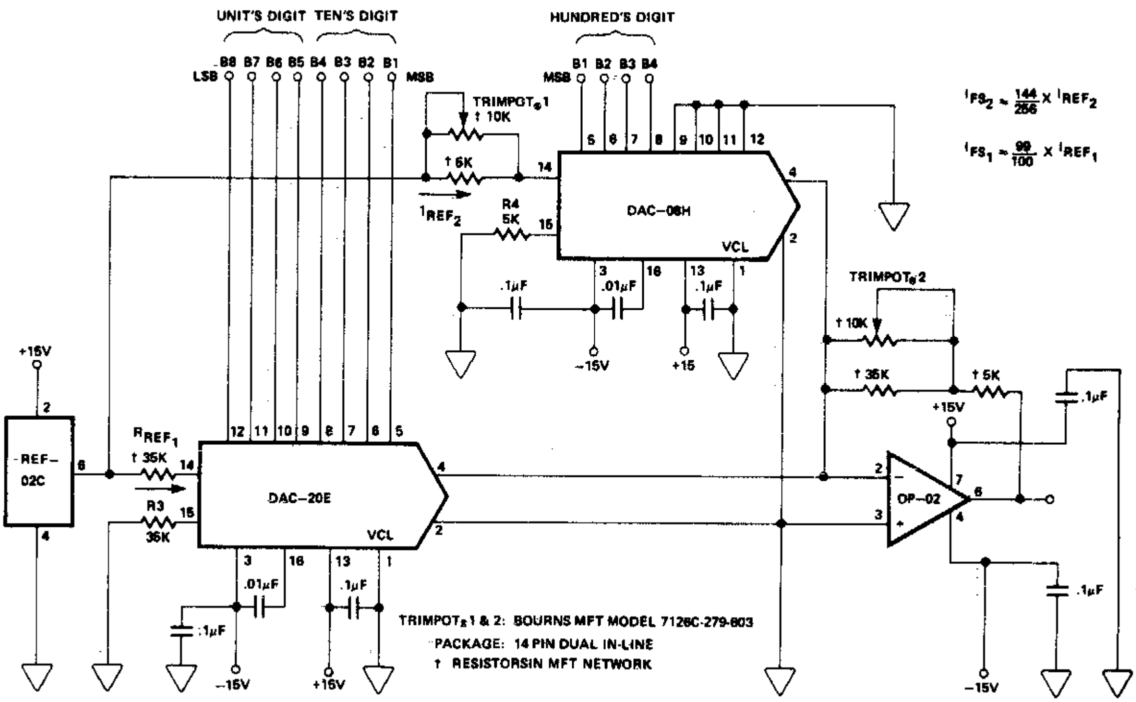


Fig. 26-5

8-BIT D/A CONVERTER

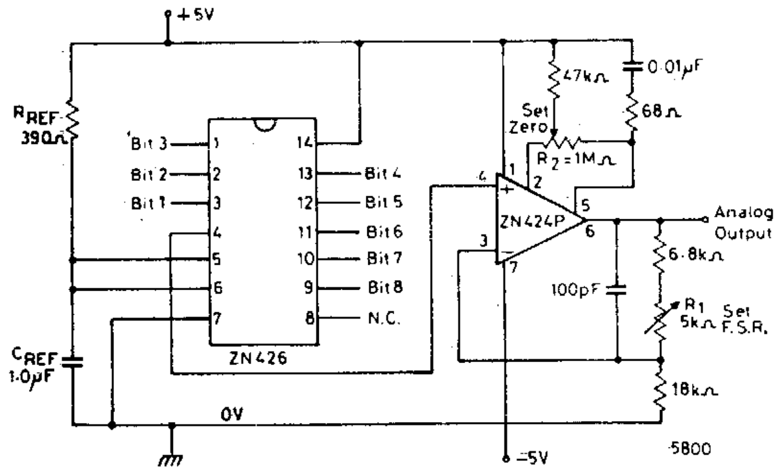


Fig. 26-6

HIGH-SPEED 8-BIT D/A CONVERTER

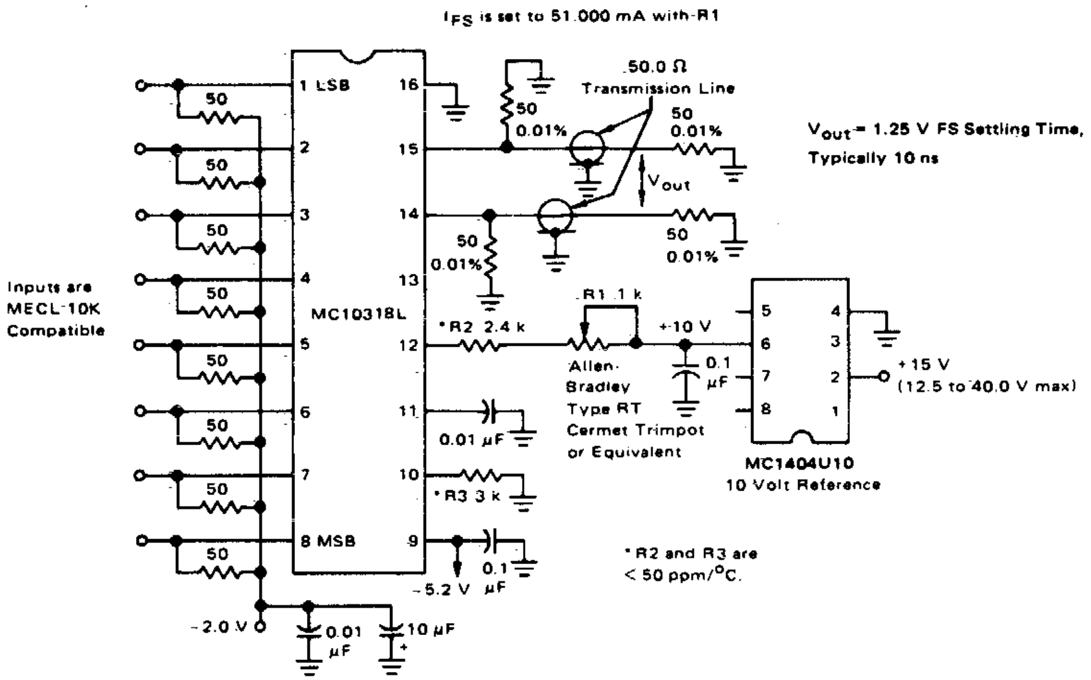
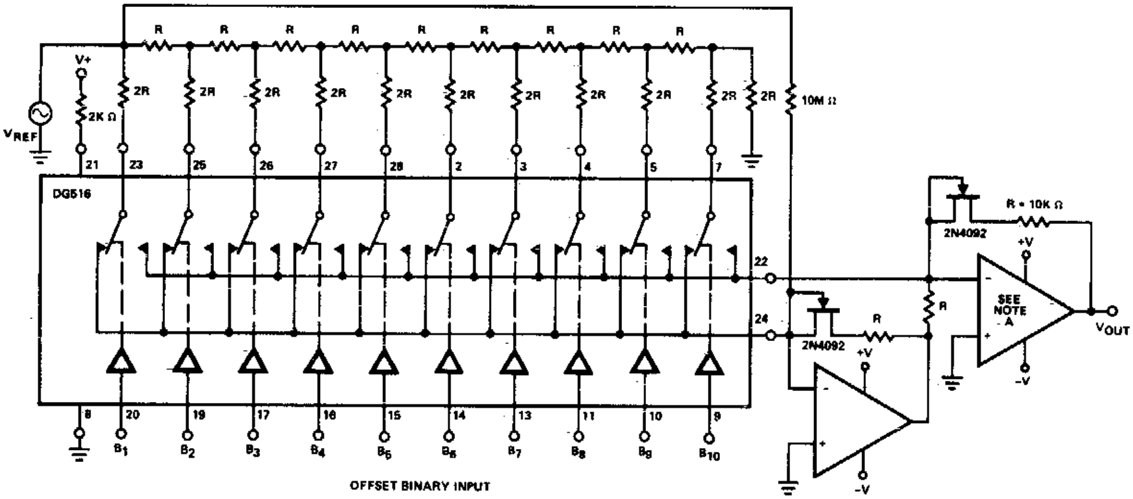


Fig. 26-7

10-BIT, 4 QUADRANT MULTIPLEXING D/A CONVERTER (OFFSET BINARY CODING)



Bipolar (Offset Binary)* Operation

DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1	$-V_{REF}\ (1-2^{-9})$
1 0 0 0 0 0 0 0 0 1	$-V_{REF}\ (2^{-9})$
1 0 0 0 0 0 0 0 0 0	0
0 1 1 1 1 1 1 1 1 1	$V_{REF}\ (2^{-9})$
0 0 0 0 0 0 0 0 0 1	$V_{REF}\ (1-2^{-9})$
0 0 0 0 0 0 0 0 0 0	V_{REF}

NOTE: 1 LSB = $2^{-9}\ V_{REF}$

*Complementing B_1 (MSB) will give 2's complement coding.

Fig. 26-8

8-BIT D/A CONVERTER

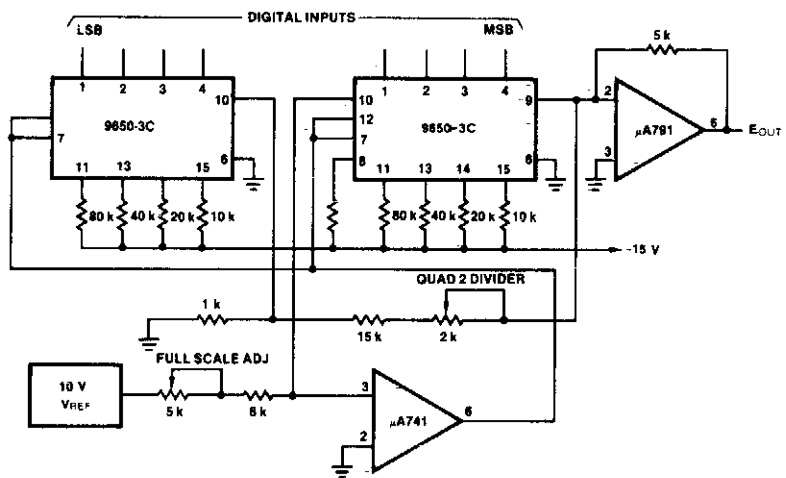
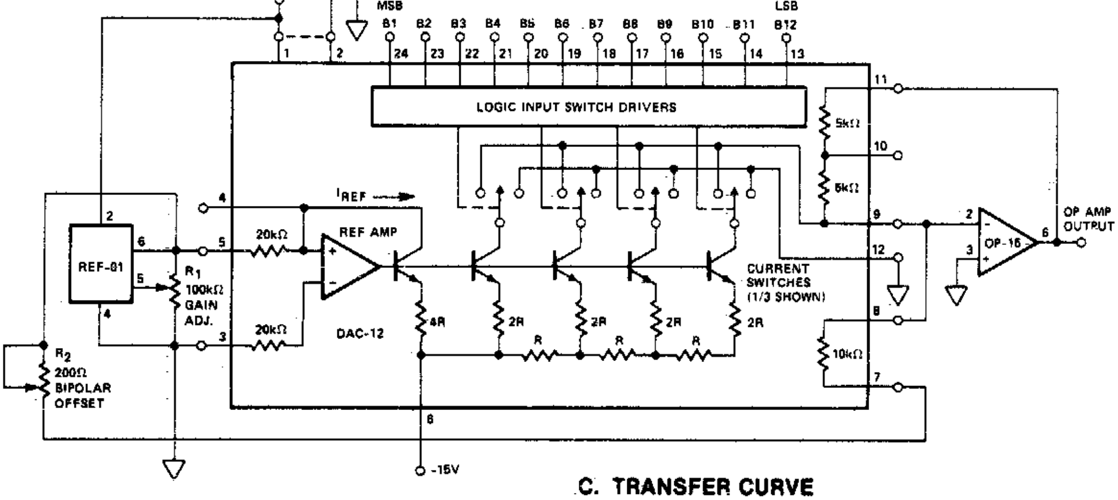


Fig. 26-9

±10 V FULL-SCALE BIPOLAR DAC

A. CONNECTION



B. ADJUSTMENT PROCEDURE

1. SET ALL BITS TO "0" STATE
2. ADJUST R_2 UNTIL OUTPUT IS −10V
3. BIT 1 to "1" STATE
4. ADJUST R_1 UNTIL OUTPUT = 0.00V

C. TRANSFER CURVE

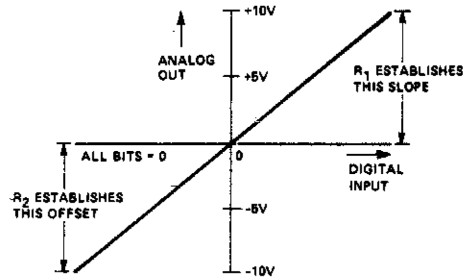


Fig. 26-10

PRECISION 12-BIT D/A CONVERTER

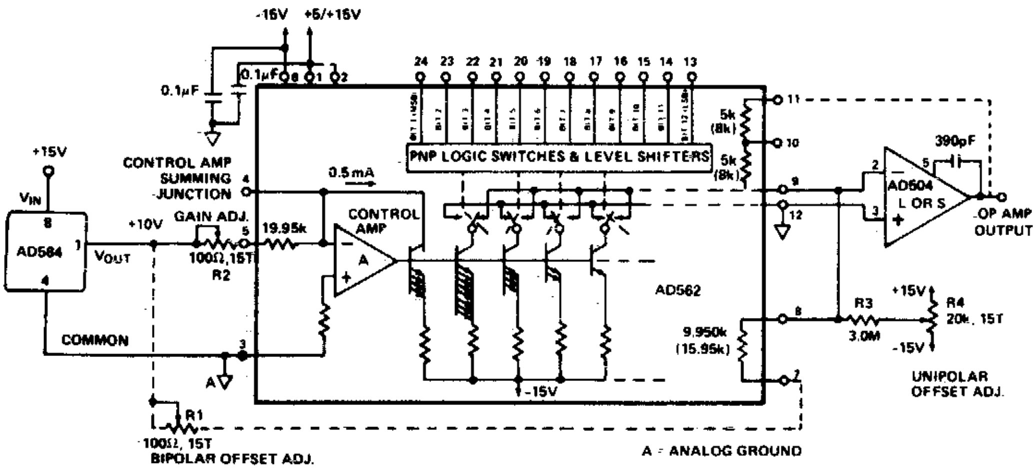


Fig. 26-11

8-BIT D/A WITH OUTPUT CURRENT-TO-VOLTAGE CONVERSION

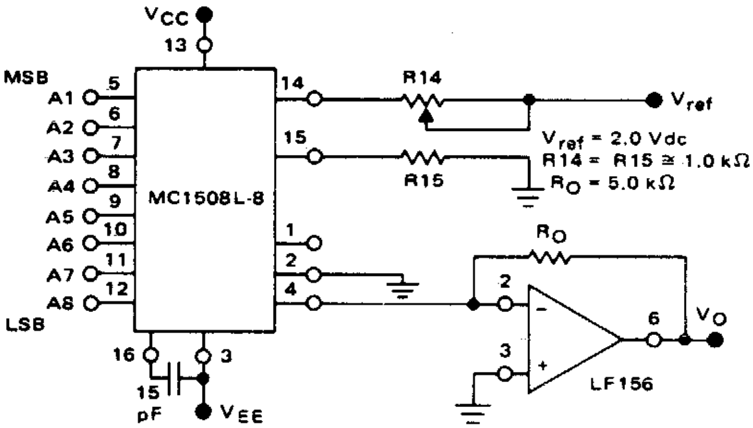


Fig. 26-12

Theoretical V_O

$$V_O = \frac{V_{ref}}{R14}(R_O)\left[\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \frac{A4}{16} + \frac{A5}{32} + \frac{A6}{64} + \frac{A7}{128} + \frac{A8}{256}\right]$$

Adjust V_{ref}, R14 or R_O so that V_O with all digital inputs at high level is equal to 9.961 volts.

$$V_O = \frac{2V}{1\,k}(5\,k)\left[\frac{1}{2} + \frac{1}{4} + \frac{1}{8} + \frac{1}{16} + \frac{1}{32} + \frac{1}{64} + \frac{1}{128} + \frac{1}{256}\right]$$

$$= 10\text{ V}\left[\frac{255}{256}\right] = 9.961\text{ V}$$

16-BIT BINARY DAC

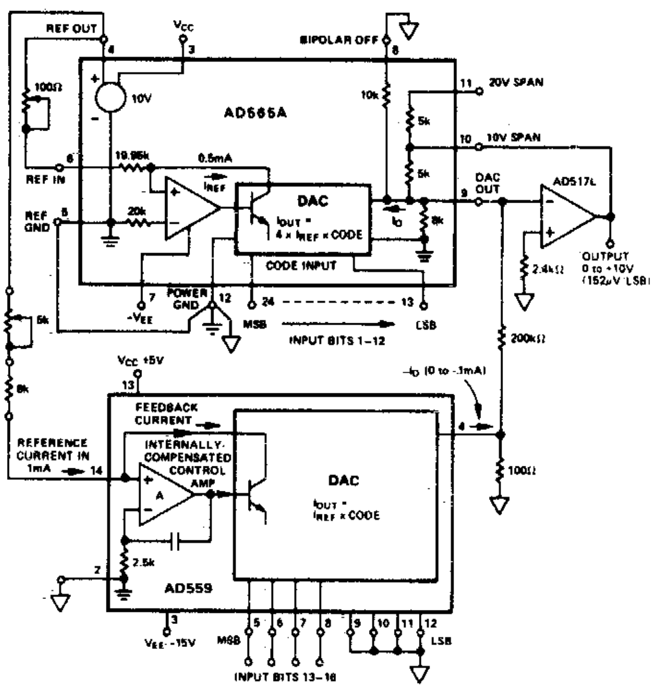
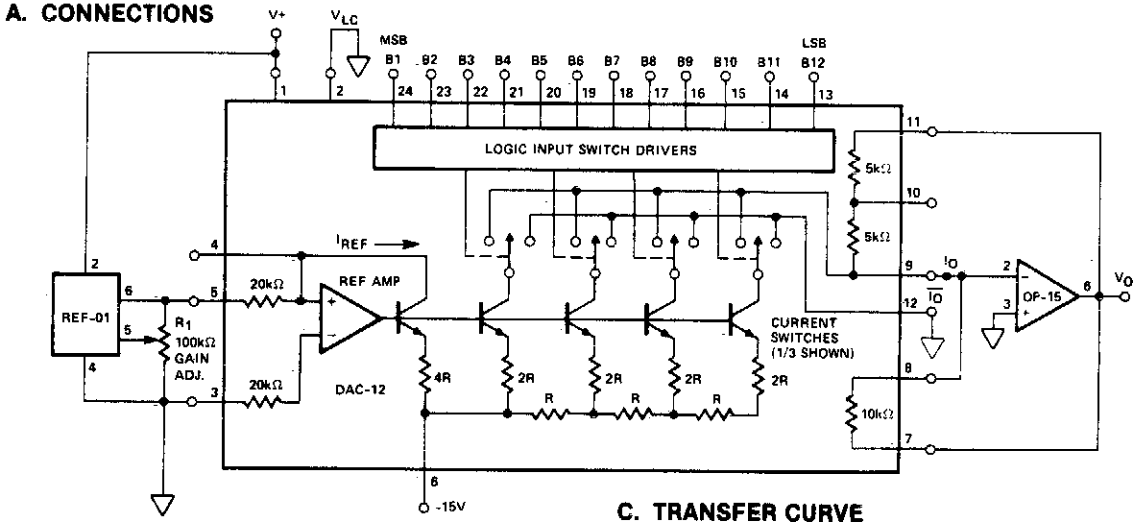


Fig. 26-13

±10 V FULL-SCALE UNIPOLAR DAC

A. CONNECTIONS



B. ADJUSTMENT PROCEDURE

1. ALL BITS TO "1" STATE ("0" STATE IF PINS 9 AND 12 INTERCHANGED)
2. ADJUST R_1 UNTIL OUTPUT IS +9.9975

$$\frac{4095}{4096} \times 10V$$

C. TRANSFER CURVE

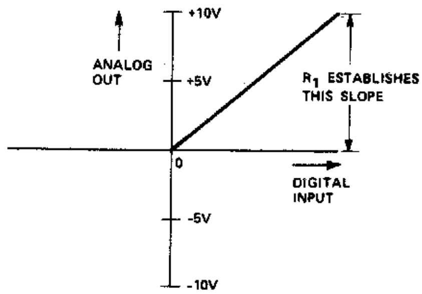
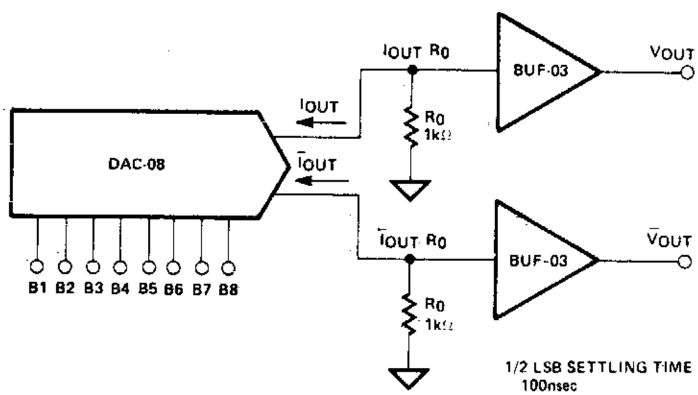


Fig. 26-14

HIGH-SPEED VOLTAGE OUTPUT DAC



SYSTEM WILL DRIVE CABLES OR TWISTED PAIRS.

Fig. 26-15

27

Dip Meters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DIP METER USING DUAL-GATE IGFET (MOSFET)

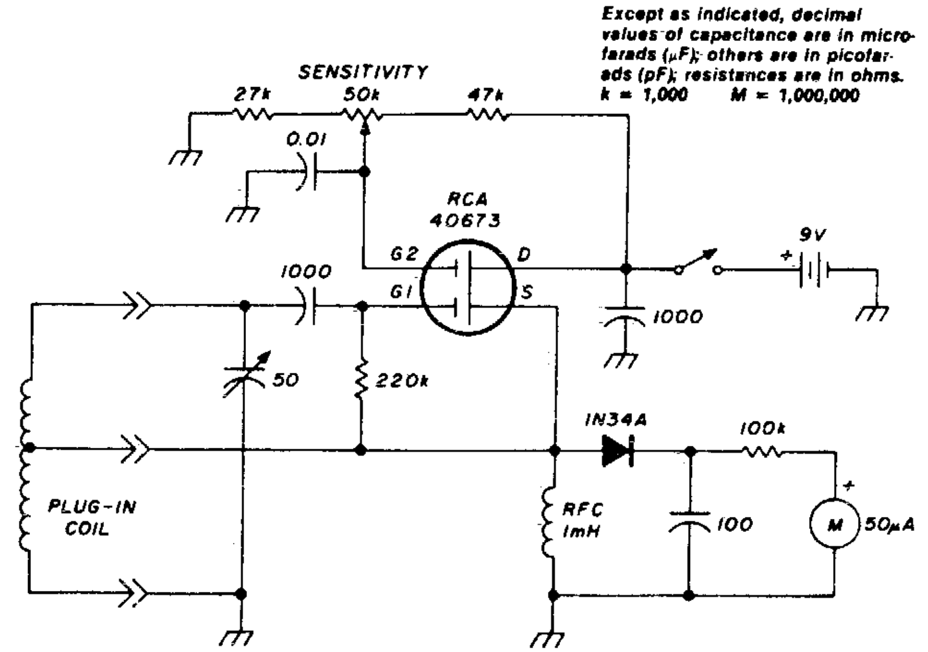
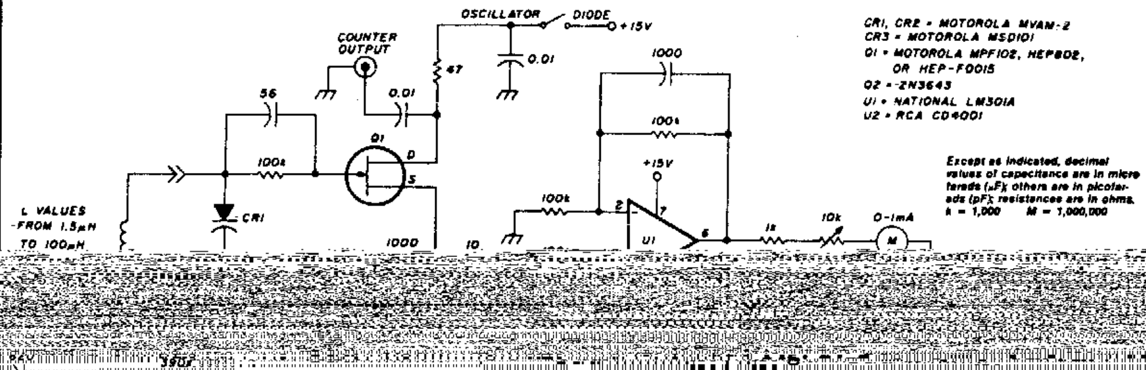


Fig. 27-1

Circuit Notes

Gate 2 is used to adjust the oscillation level.

VARICAP-TUNED FET DIP METER WITH 1-kHz MODULATION



DIP METER USING N-CHANNEL IGFET (MOSFET) AND SEPARATE DIODE DETECTOR

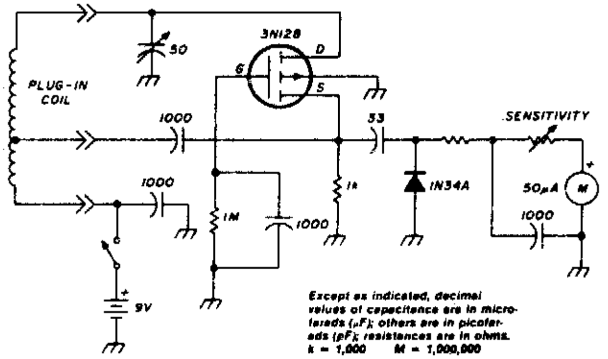
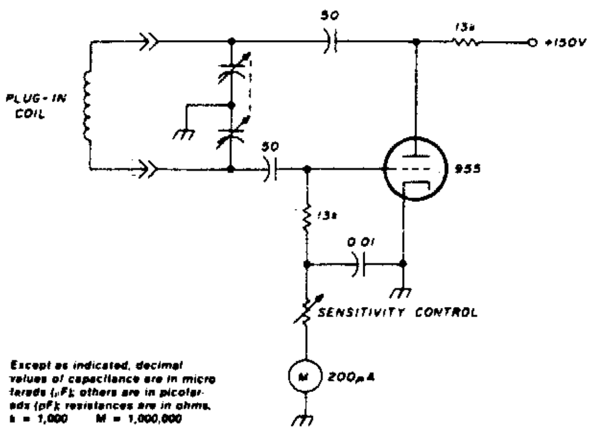


Fig. 27-3

BASIC GRID-DIP METER



Circuit Notes

This circuit uses a triode vacuum-tube (9002 and 6C4 also commonly used).

Fig. 27-4

DIP METER USING GERMANIUM PNP BIPOLAR TRANSISTOR WITH SEPARATE DIODE DETECTOR

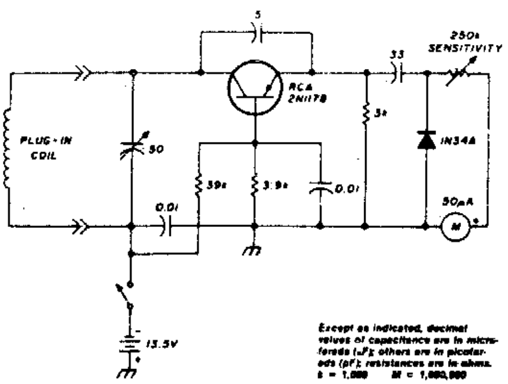
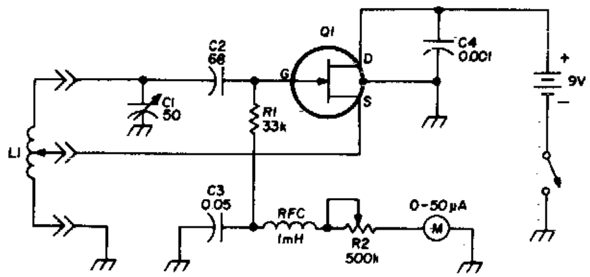


Fig. 27-5

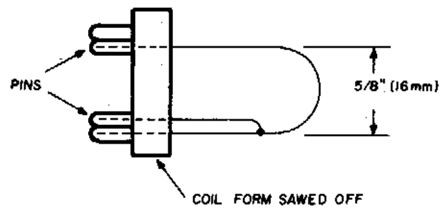
GATE-DIP METER COVERS 1.8 - 150 MHz



Coil data.

frequency range (MHz)	no. turns	wire size AWG	wire size (mm)	winding length inches	winding length (mm)	tap*	coil diameter inches	coil diameter (mm)
1.8 - 3.8	82	26 enamel	(0.4)	1 9/16	(40.0)	12	1¼	(32)
3.6 - 7.3	29	26 enamel	(0.4)	9/16	(14.5)	5	1¼	(32)
7.3 - 14.4	18	22 enamel	(0.6)	3/4	(19.0)	3	1	(25)
14.4 - 32	7	22 enamel	(0.6)	1/2	(12.5)	2	1	(25)
29 - 64	3½	18 tinned	(1.0)	3/4	(19.0)	3/4	1	(25)

61 - 150 Hairpin of 16 no. AWG (1.3mm) wire, 5/8 inch (16mm) spacing, 2 3/8 inches (60mm) long including coil-form pins. Tapped at 2 inches (51mm) from ground end.



*Turns from ground-end. 1 inch (25mm) forms are Millen 45004 available from Burstein-Applebee

Fig. 27-6

DIP METER USING SILICON JUNCTION FET

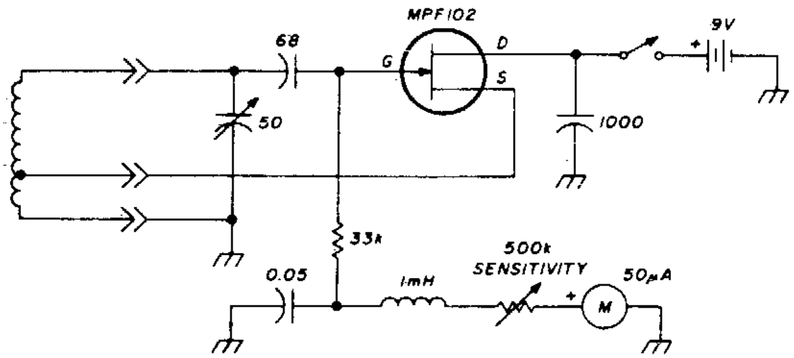


Fig. 27-7

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

28

Displays

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LED Brightness Control
LED Bar/Dot Level Meter
60 dB Dot Mode Display
Bar Display with Alarm Flasher
12-Hour Clock with Gas Discharge Displays
Precision Frequency Counter (~ 1 MHz Maximum)
Exclamation Point Display
LED Bar Peak Program Meter Display for Audio
10 MHz Universal Counter

LED BRIGHTNESS CONTROL

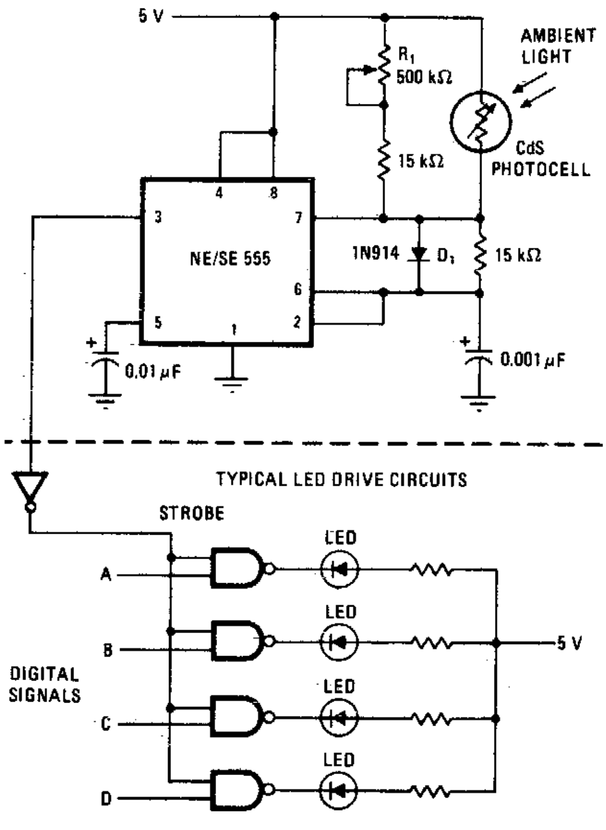


Fig. 28-1

Circuit Notes

The brightness of LED display is varied by using a photocell in place of one timing resistor in a 555 timer, and bypassing the other timing resistor to boost the timer's maximum duty cycle. The result is a brighter display in sunlight and a fainter one in the dark.

LED BAR/DOT LEVEL METER

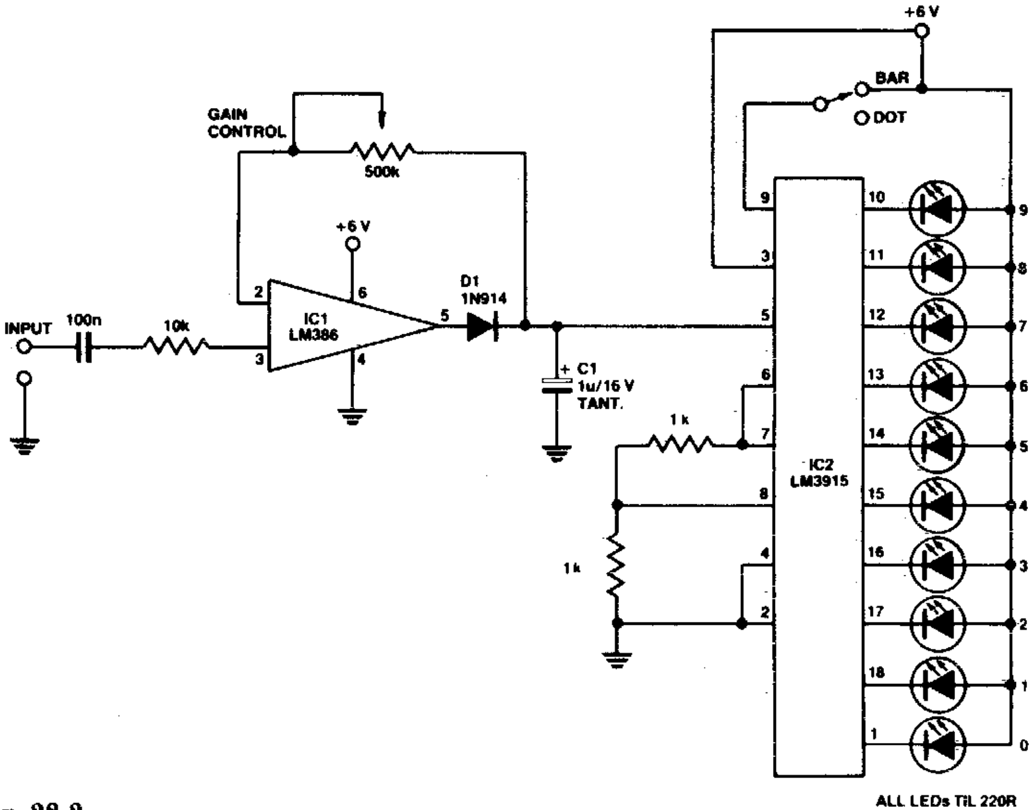
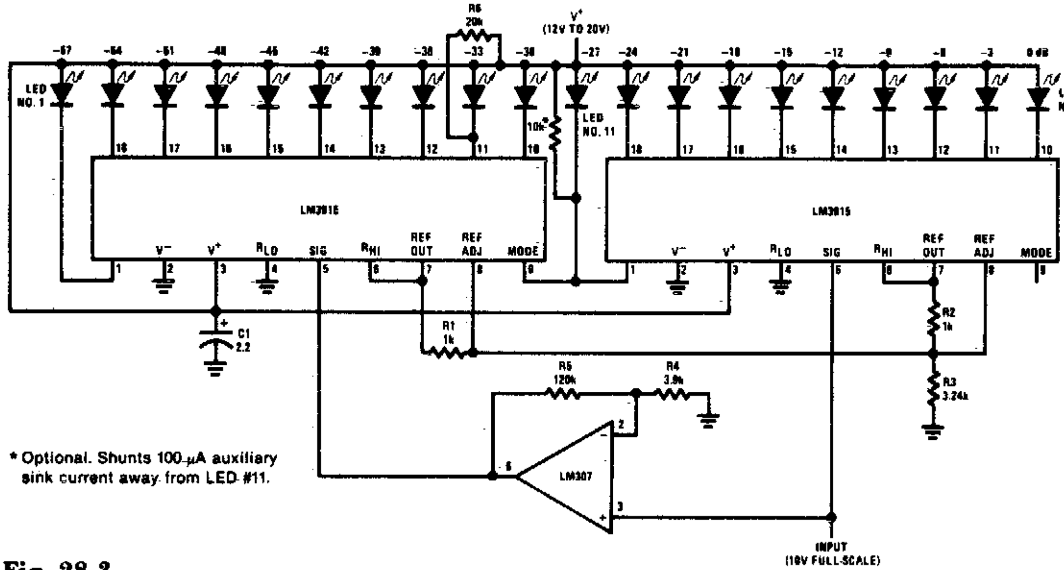


Fig. 28-2

Circuit Notes

A simple level of power meter can be arranged to give a bar or dot display for a hi-fi system. Use green LEDs for 0 to 7; yellow for 8 and red for 9 to indicate peak power. The gain control is provided to enable calibration on the equipment with which the unit is used. Because the unit draws some 200 mA, a power supply is advisable instead of running the unit from batteries.

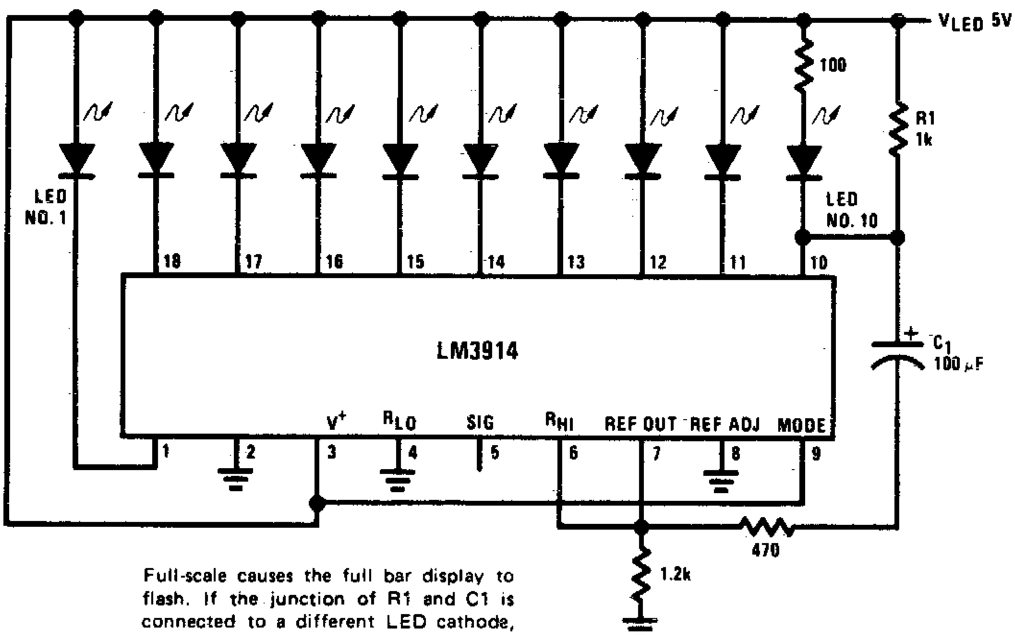
60 dB DOT MODE DISPLAY



* Optional. Shunts 100 μA auxiliary sink current away from LED #11.

Fig. 28-3

BAR DISPLAY WITH ALARM FLASHER



Full-scale causes the full bar display to flash. If the junction of R1 and C1 is connected to a different LED cathode, the display will flash when that LED lights, and at any higher input signal.

Fig. 28-4

12-HOUR CLOCK WITH GAS DISCHARGE DISPLAYS

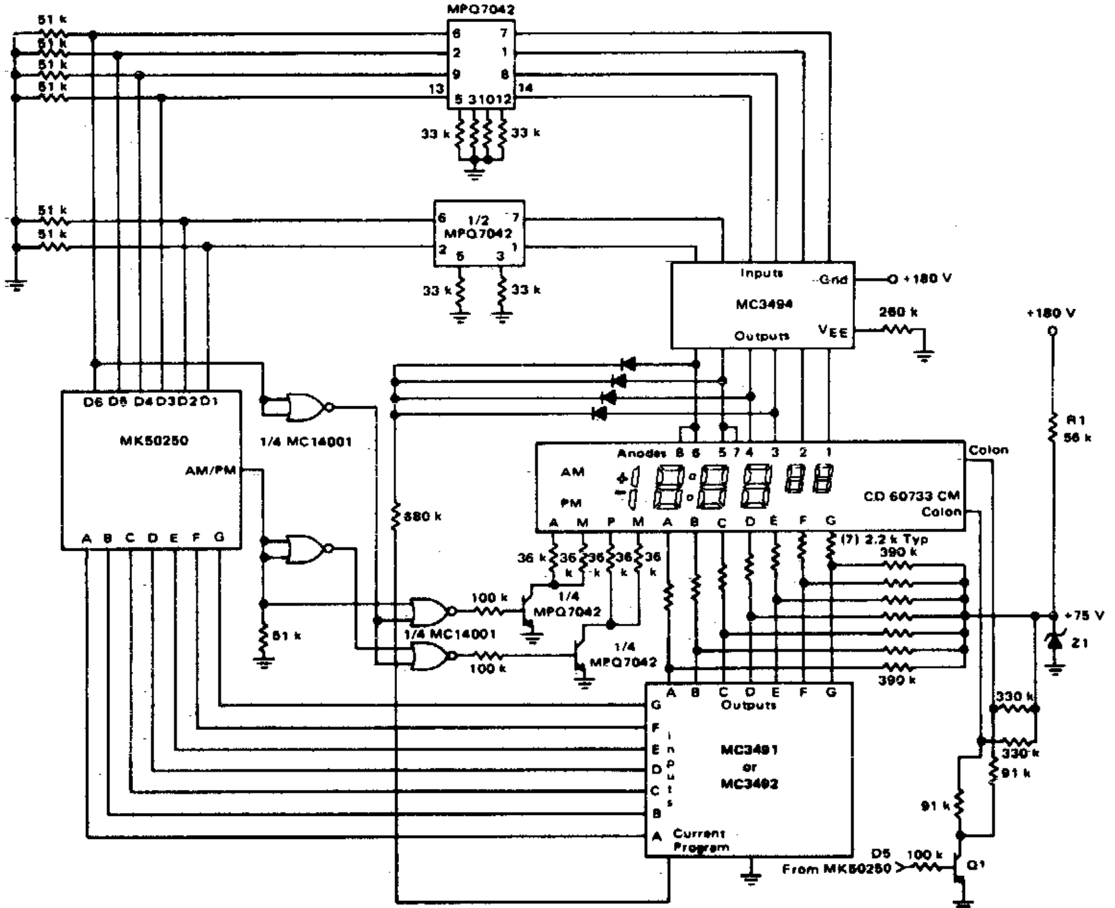


Fig. 28-5

PRECISION FREQUENCY COUNTER (~1 MHz MAXIMUM)

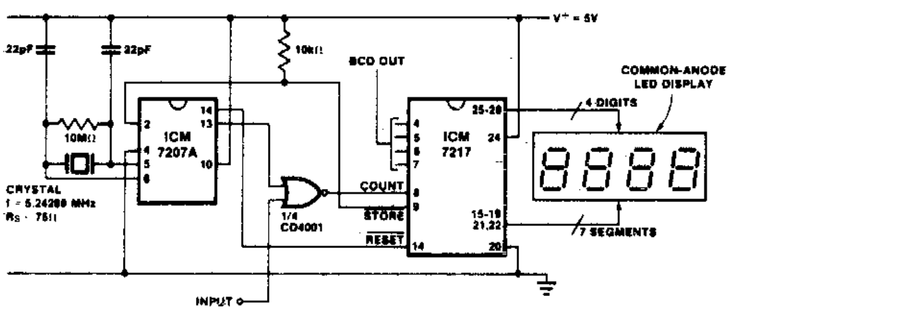
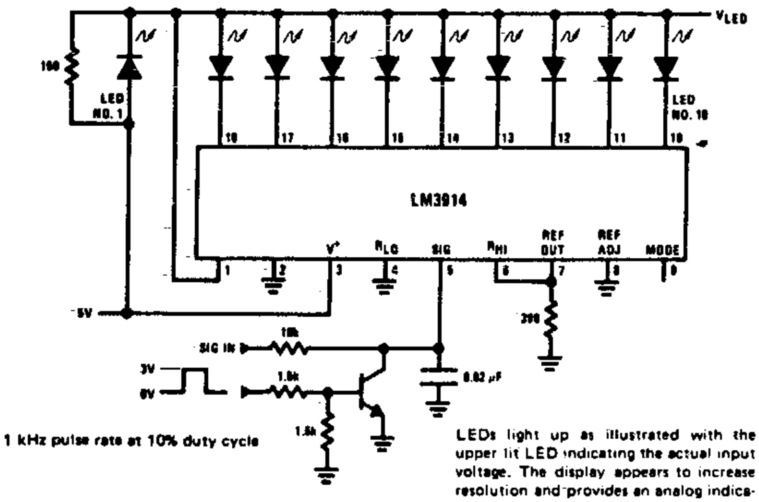


Fig. 28-6

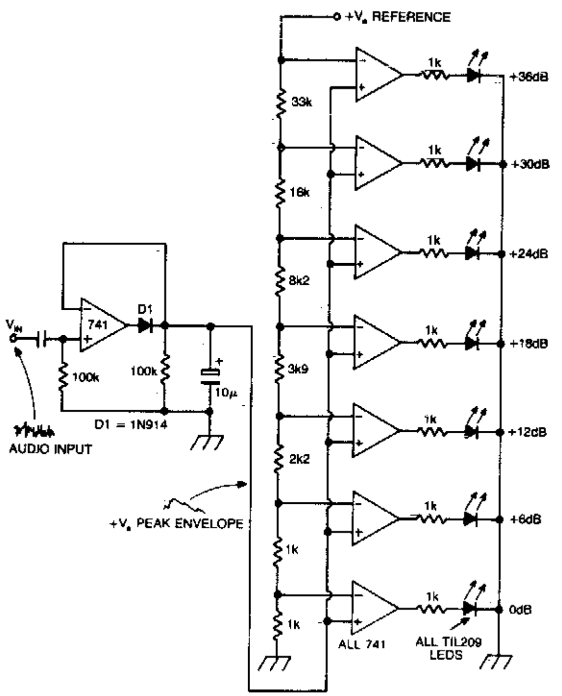
EXCLAMATION POINT DISPLAY



LEDs light up as illustrated with the upper lit LED indicating the actual input voltage. The display appears to increase resolution and provides an analog indication of overrange.

Fig. 28-7

LED BAR PEAK PROGRAM METER DISPLAY FOR AUDIO



Circuit Notes

A bar column of LEDs is arranged so that as the audio signal level increases, more LEDs in the column light up. The LEDs are arranged vertically in 6 dB steps. A fast response time and a one second decay time give an accurate response to transients and a low "flicker" decay characteristic. On each of the op amps inverting inputs is a dc reference voltage, which increases in 6 dB steps. All noninverting inputs are tied together and connected to the positive peak envelope of the audio signal. Thus, as this envelope exceeds a particular voltage reference, the op amp output goes high and the LED lights up. Also, all the LEDs below this are illuminated.

Fig. 28-8

10 MHz UNIVERSAL COUNTER

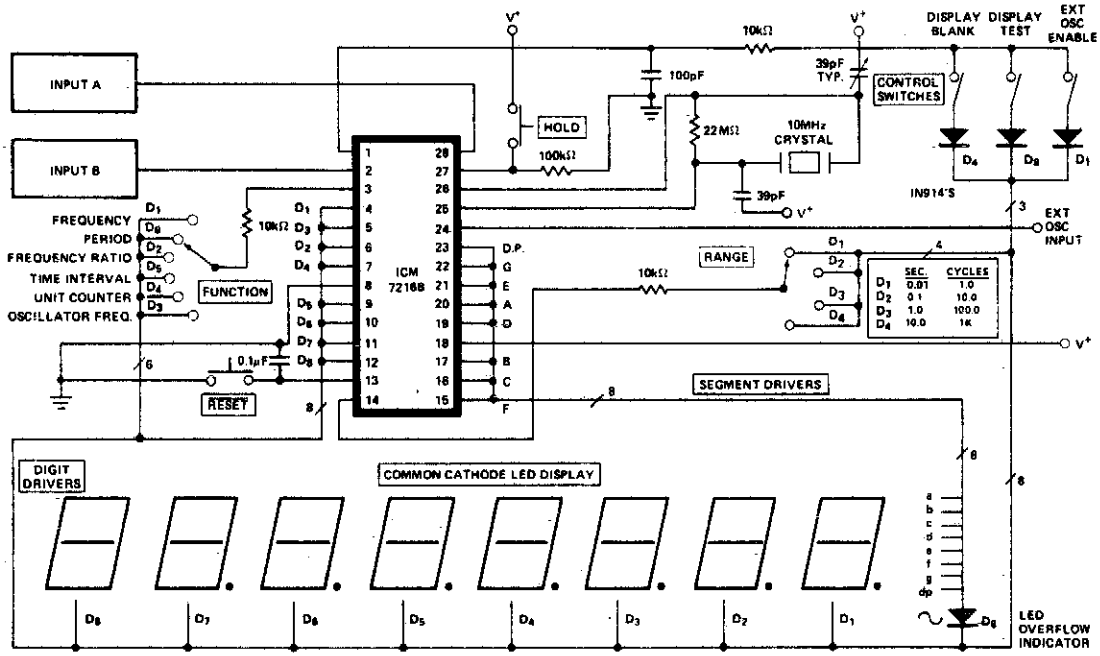


Fig. 28-9

Circuit Notes

This is a minimum component complete Universal Counter. It can use input frequencies up to 10 MHz at INPUT A and 2 MHz at INPUT B. If the signal at INPUT A has a very low duty cycle, it may be necessary to use a 74121 monostable multivibrator or similar circuit to stretch the input pulse width to be able to guarantee that it is at least 50 ns in duration.

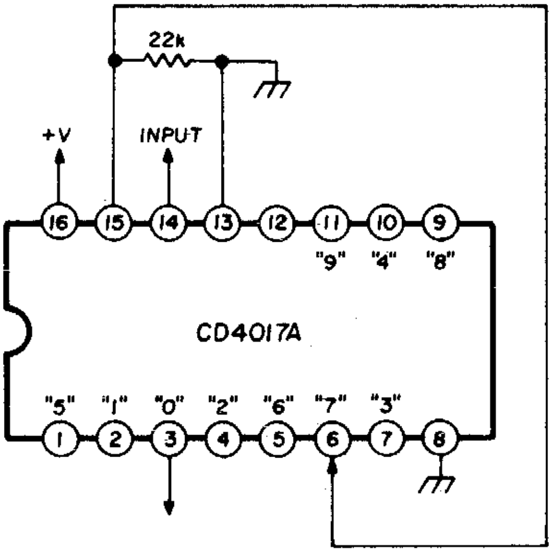
29

Dividers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

CMOS Programmable Divide-by-N Counter
Frequency Divider Chain
Frequency Divider with Transient
Free Output
Binary Divider Chain
Decade Frequency Divider

CMOS PROGRAMMABLE DIVIDE-BY-N COUNTER



Circuit Notes

A single connection change permits division by any integer between 2 and 10. The RCA CD4017A Johnson decade counter is shown connected as a divide by 7 counter. The resistor is used to hold the reset line low. When the appropriate number is reached, that output and the reset line are driven high, resetting the counter. To divide by other integers, pin 15 should be connected to the desired output. For example, pin 1 for a divide by 5, or pin 7 for a divide by 3. The output of the divider appears on the 0 line.

Fig. 29-1

FREQUENCY DIVIDER CHAIN

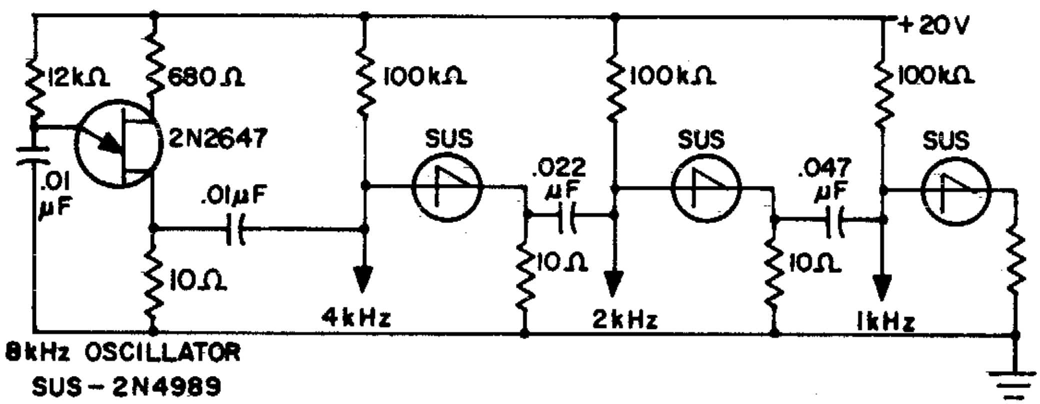


Fig. 29-2

Circuit Notes

Sawtooth output from each stage is one half frequency of preceding stage.

FREQUENCY DIVIDER WITH TRANSIENT FREE OUTPUT

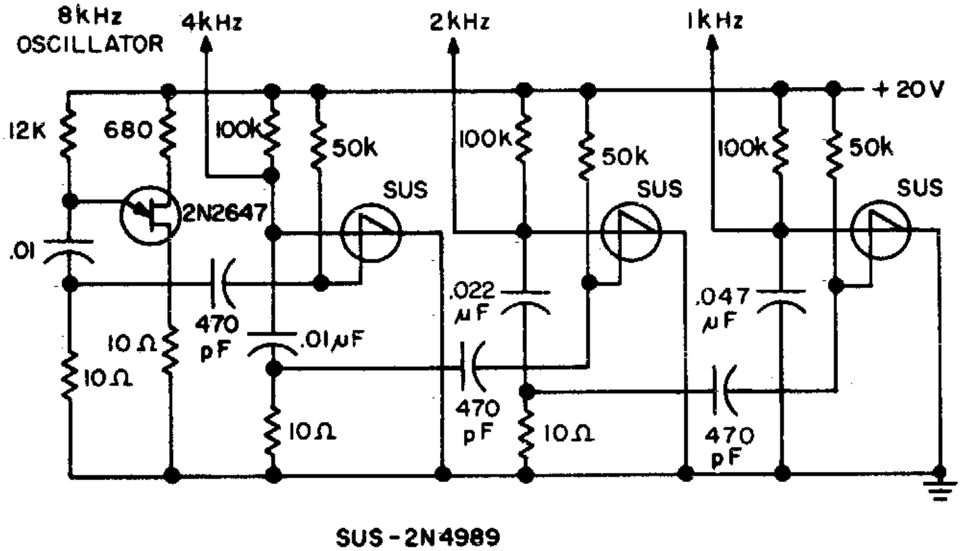


Fig. 29-3

Circuit Notes

Spikes in the center of a sawtooth wave are eliminated in this circuit by triggering at gate.

BINARY DIVIDER CHAIN

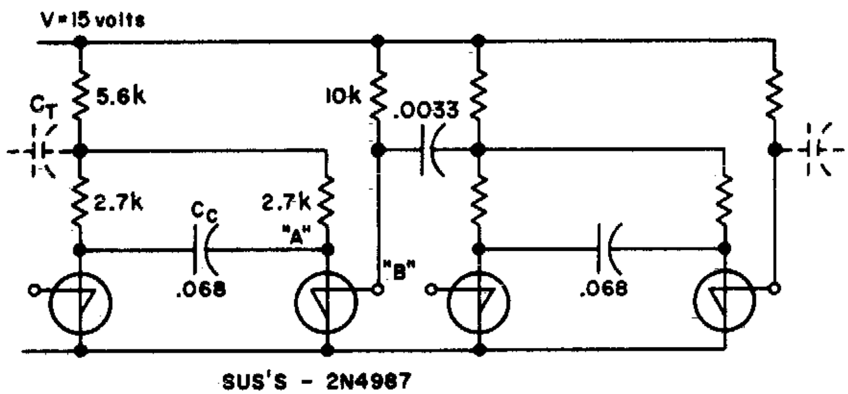


Fig. 29-4

Circuit Notes

This circuit uses fewer components than transistor flip flops. Output at "B" gives a transient-free waveform.

DECADE FREQUENCY DIVIDER

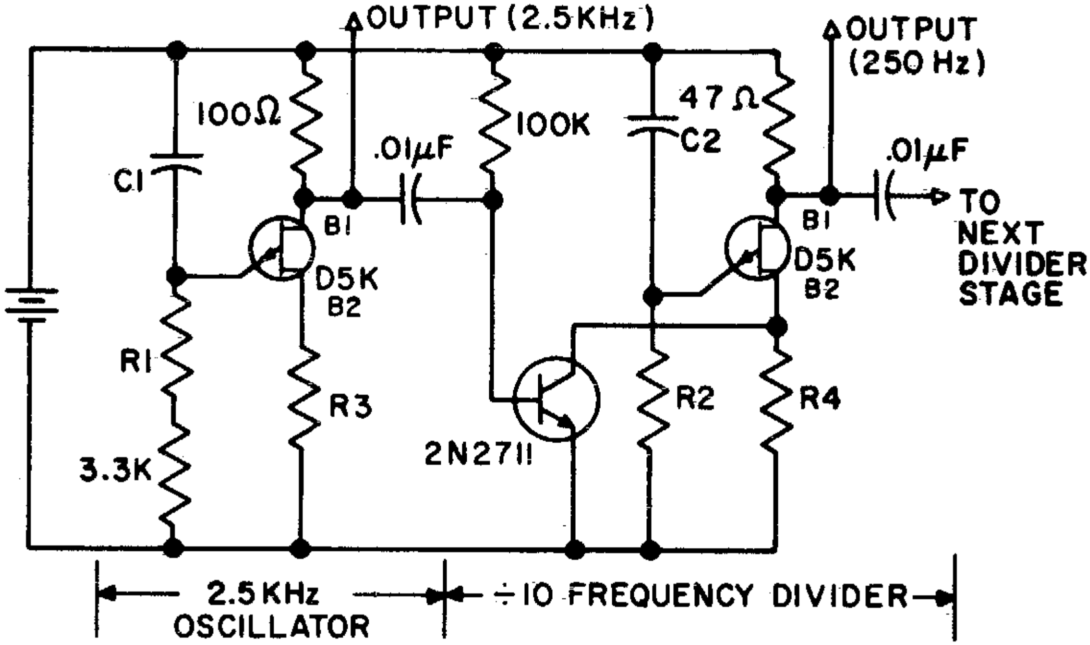


Fig. 29-5

Circuit Notes

In the next stage, the product of R2 and C2 should be 10 × that of the preceding stage (±2%). R2 should be between 27K and 10 M.

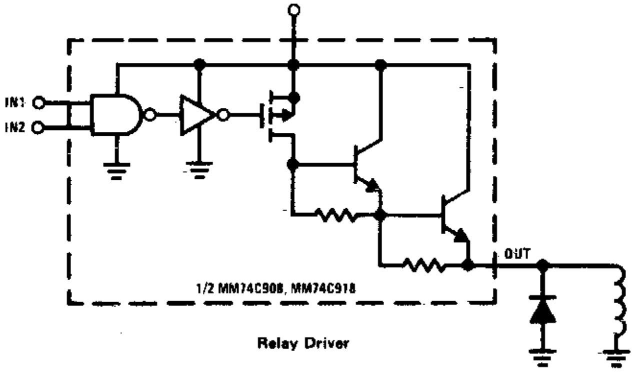
C1 & C2—.0047 μF (±1 %)
R1—100K (±1%)
R2—1M (±1%)
R3—R4—1K (may need to be adjusted for variation of R_{BB} of UJT)

30

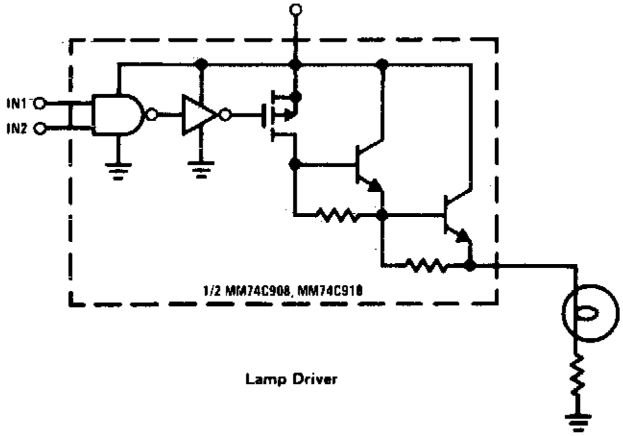
Drivers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DRIVER CIRCUITS

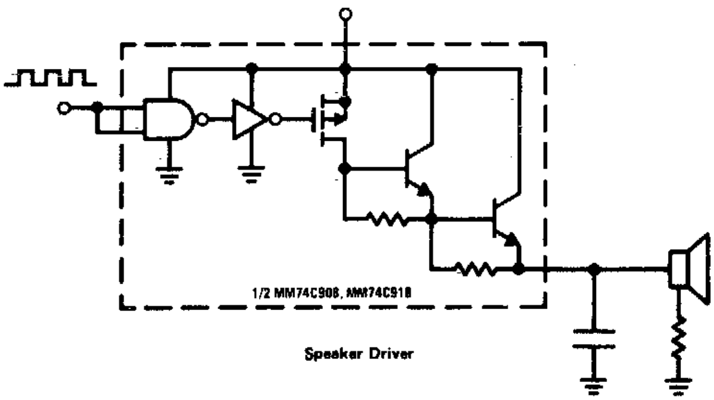


Relay Driver



Lamp Driver

Fig. 30-1

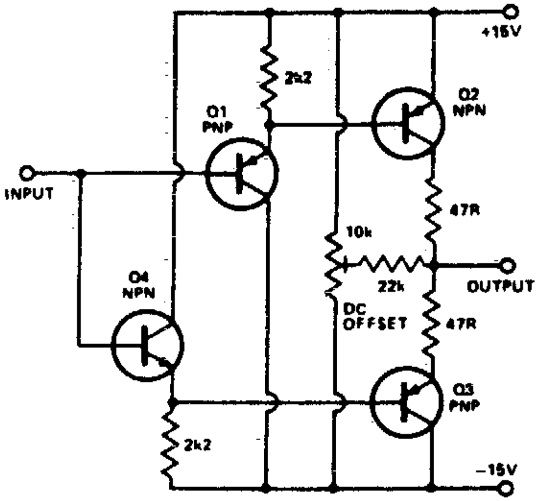


Speaker Driver

Circuit Notes

CMOS drivers for relays, lamps, speakers, etc., offers extremely low standby power. At $V_{CC} = 15$ V, power dissipation per package is typically 750 nW when the outputs are not drawing current. Thus, the drivers can be sitting out on line (a telephone line, for example) drawing essentially zero current until activated.

50 OHM DRIVER



Circuit Notes

To buffer a test generator to the outside world requires an amplifier with sufficient bandwidth and power handling capability. The circuit is a very simple unity gain buffer. It has a fairly high input impedance, a 50 ohm output impedance, a wide bandwidth, and high slew rate. The circuit is simply two pairs of emitter followers. The base emitter voltages of Q1 and Q2 cancel out, and so do those of Q3 and Q4. The preset is used to zero out any small dc offsets due to mismatching in the transistors.

Fig. 30-2

LINE DRIVER

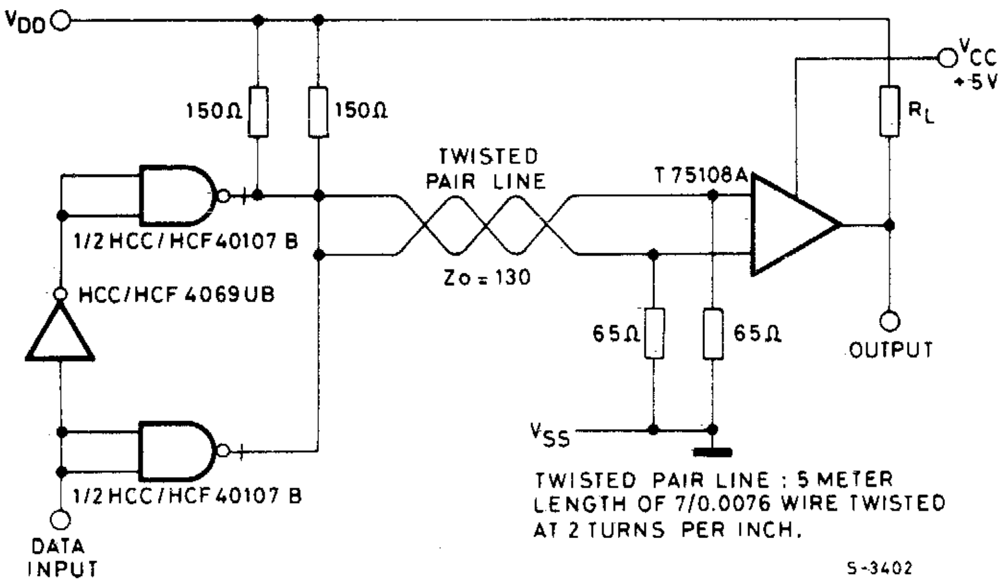


Fig. 30-3

HIGH-SPEED LASER DIODE DRIVER

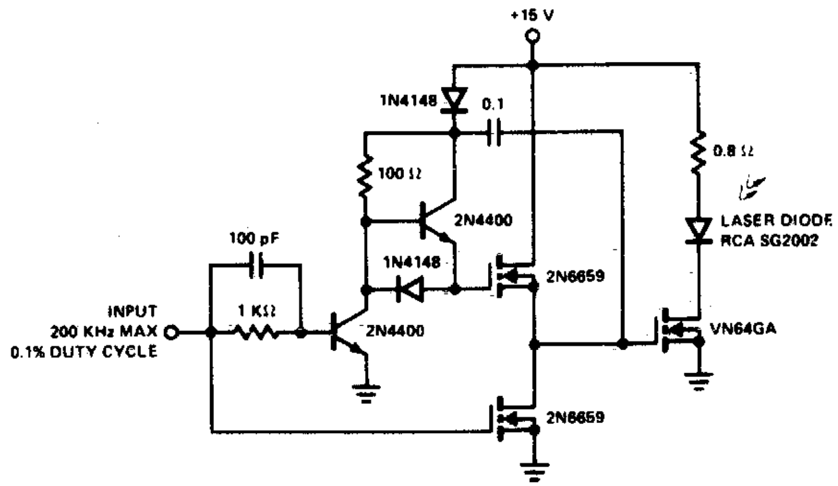


Fig. 30-4

Circuit Notes

A faster driver can supply higher peak gate current to switch the VN64GA very quickly. The circuit uses a VMOS totempole stage to drive the high power switch.

CAPACITIVE LOAD DRIVER

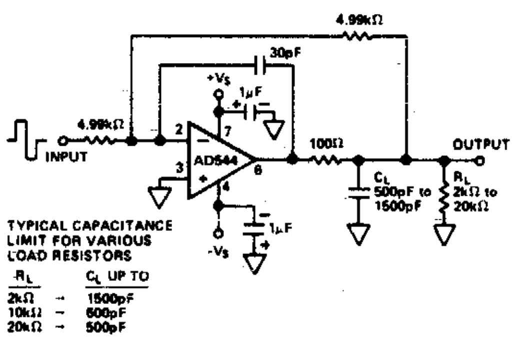


Fig. 30-5

Circuit Notes

The circuit employs a 100 ohm isolation resistor which enables the amplifier to drive capacitive loads exceeding 500 pF; the resistor effectively isolates the high frequency feedback from the load and stabilizes the circuit. Low frequency feedback is returned to the amplifier summing junction via the low pass filter formed by the 100 ohm series resistor and the load capacitance. C_L.

RELAY DRIVER

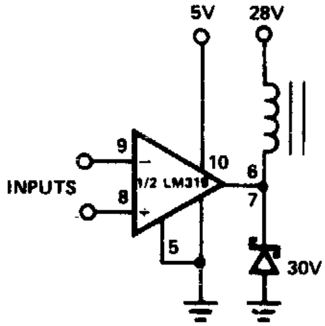
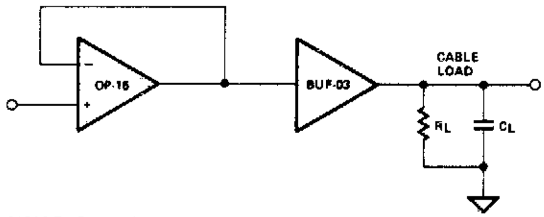


Fig. 30-6

BIFET CABLE DRIVER



CAPACITIVE LOAD STABILITY OF BUF-03 MAKES IT AN IDEAL INTERFACE BETWEEN BIFET OP AMPS AND SHIELDED CABLES'

NOTE: TO MAINTAIN ACCURACY IN THE BUFFER $R_L > 1k\Omega$ IS RECOMMENDED.

Fig. 30-8

RELAY DRIVER

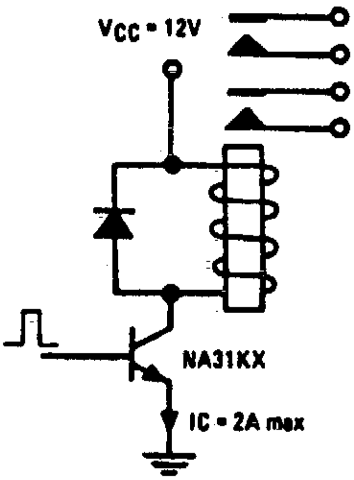
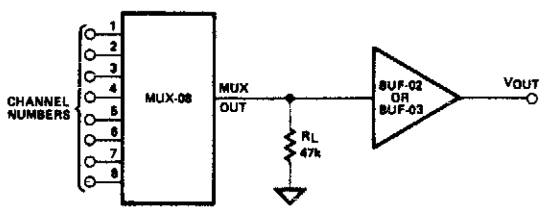


Fig. 30-7

HIGH SPEED LINE DRIVER FOR MULTIPLEXERS



NOTE 1: STRAY CAPACITANCE AT MULTIPLEXER OUTPUT NODE SHOULD BE MINIMIZED TO REDUCE CHANNEL-TO-CHANNEL CROSSTALK.

NOTE 2: A BUFFER WHOSE SLEW RATE IS TOO SMALL WILL INCREASE CHANNEL-TO-CHANNEL CROSSTALK.

Fig. 30-9

HIGH IMPEDANCE METER DRIVER

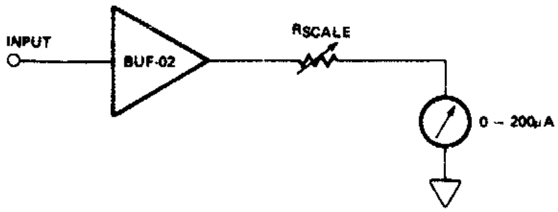


Fig. 30-10

CRT YOKE DRIVER

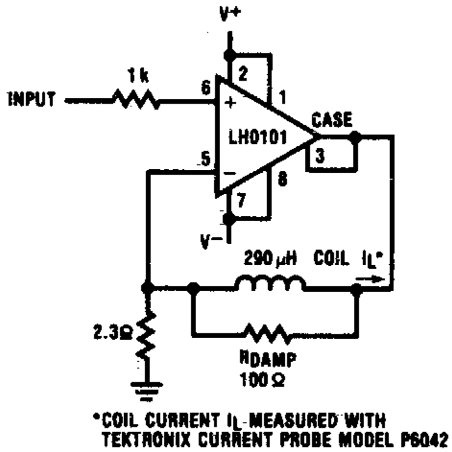


Fig. 30-12

Circuit Notes

A 500 mV peak-to-peak triangular waveform about ground is input to the amplifier, giving rise to a 100 mA peak current to the inductor.

CRT DEFLECTION YOKE DRIVER

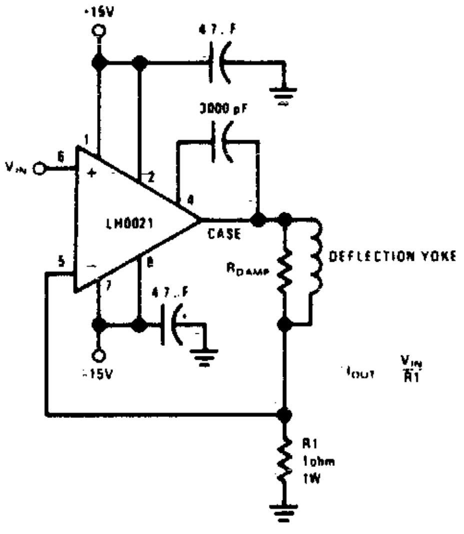


Fig. 30-11

SOLENOID DRIVER

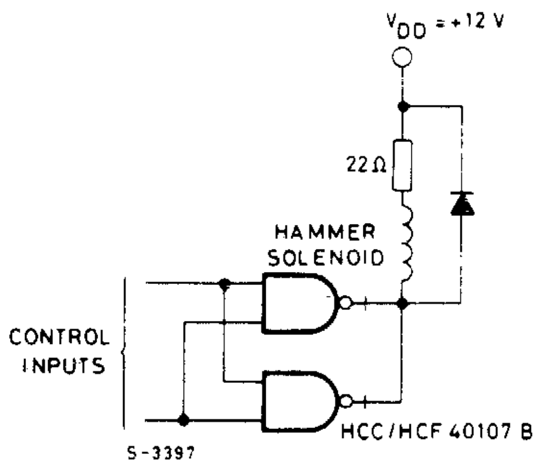


Fig. 30-13

COAXIAL CABLE DRIVER

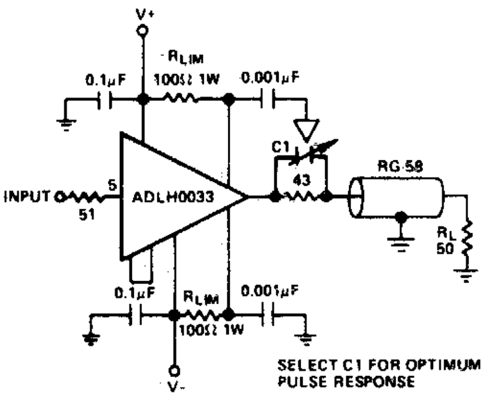


Fig. 30-14

HIGH SPEED SHIELD/LINE DRIVER

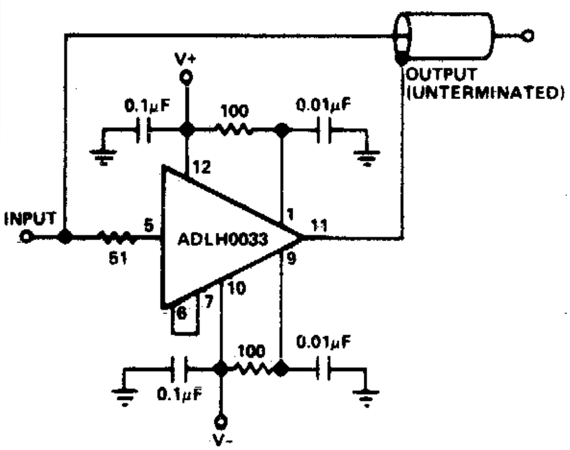
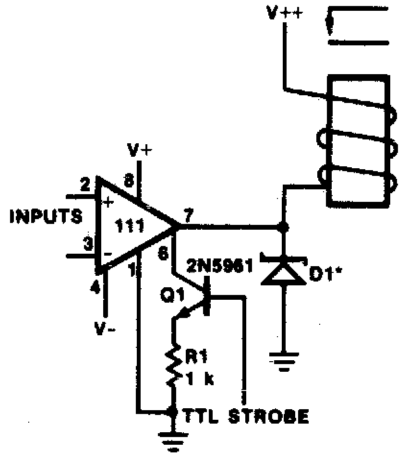


Fig. 30-15

RELAY DRIVER WITH STROBE



*Absorbs inductive kickback of relay and protects IC from severe voltage transients on V++ line.

Fig. 30-16

DIRECT DC DRIVE INTERFACE OF A TRIAC

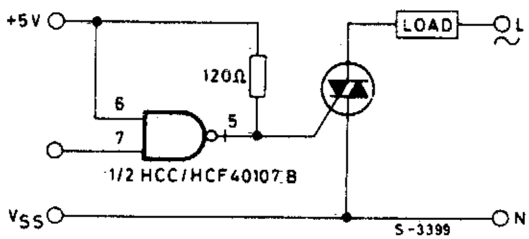


Fig. 30-17

31

Fiber Optic Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FIBER-OPTICS HALF DUPLEX INFORMATION LINK

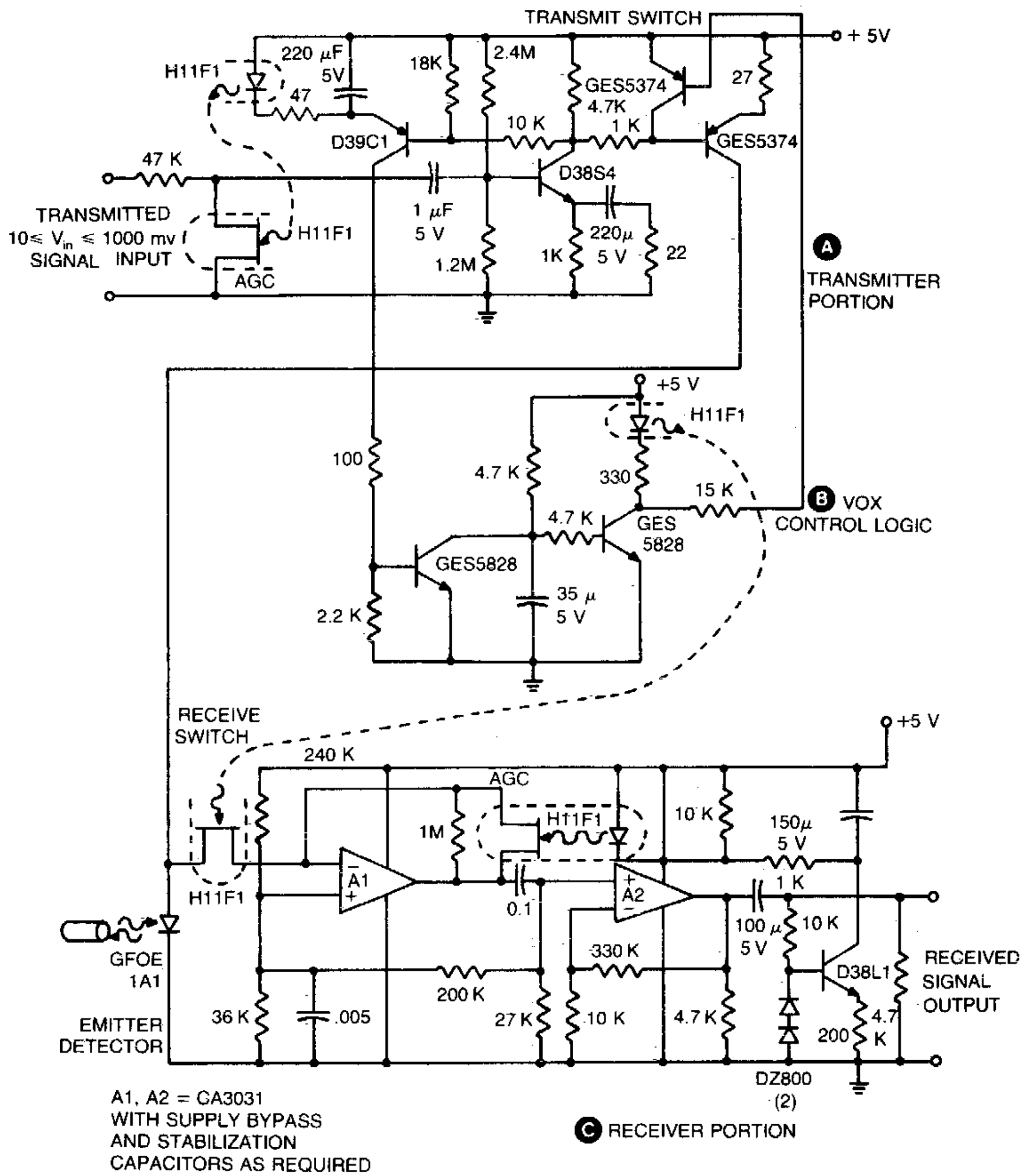


Fig. 31-1

FIBER-OPTIC RECEIVER, VERY HIGH SENSITIVITY, LOW SPEED, 3nW

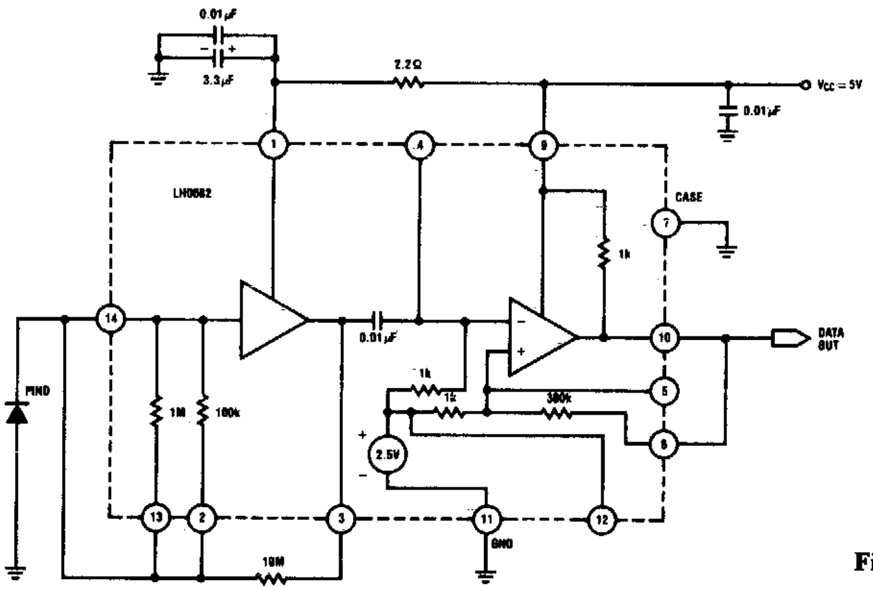


Fig. 31-2

FIBER-OPTIC LINK

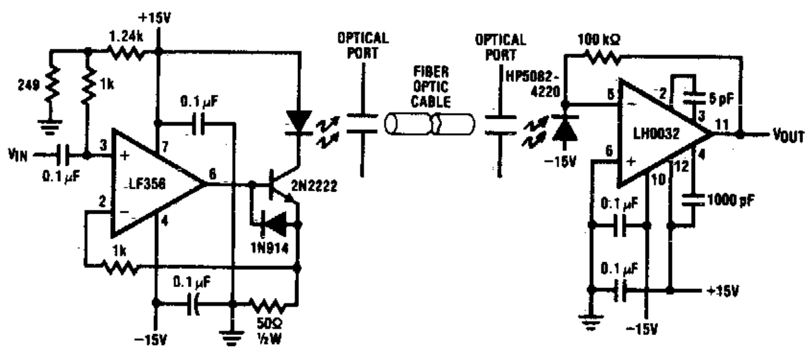


Fig. 31-3

Circuit Notes

Fiber Optic applications require analog drivers and receivers operating in the megahertz region. This complete analog transmission system is suitable for optical communication applications up to 3.5 MHz. The transmitter LED is normally biased at 50 mA operating current. The input is capacitively coupled and ranges from 0 to 5 V, modulating the LED current from 0 to 100 mA. The receiver circuit is configured as a transimpedance amplifier. The photodiode with 0.5 amp per watt responsivity generates a 50 mV signal at the receiver output for 1 μW of light input.

FIBER-OPTIC LINK REPEATER

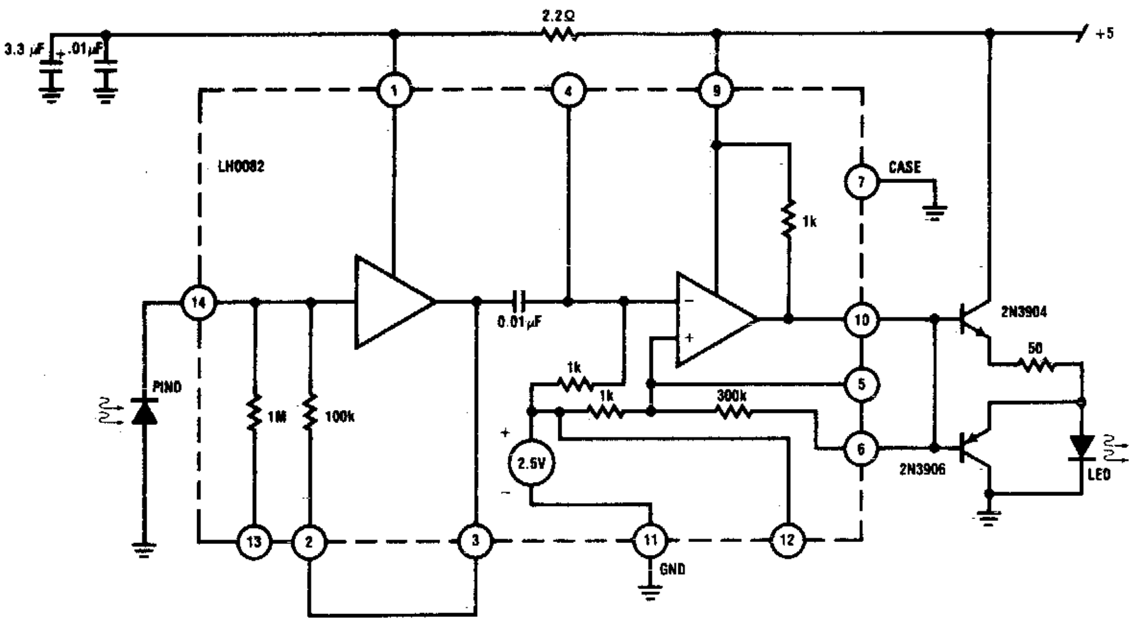


Fig. 31-4

FIBER-OPTIC RECEIVER, HIGH SENSITIVITY, 30nW

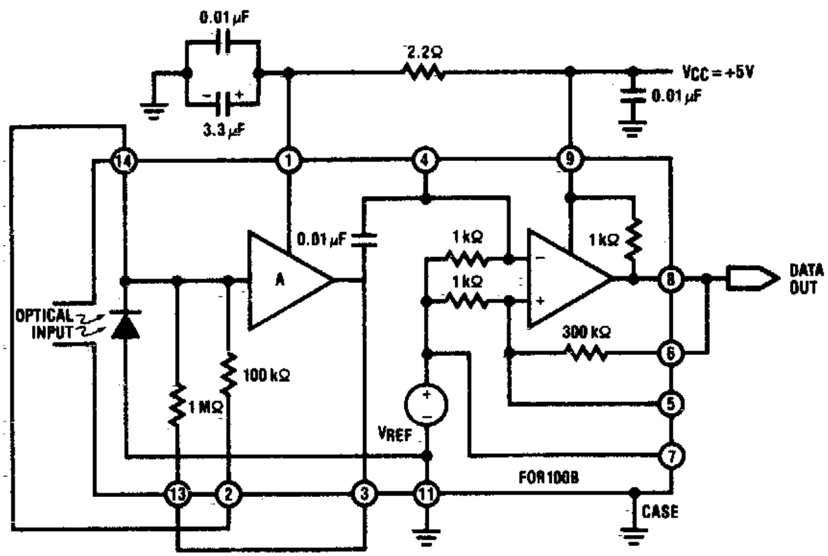


Fig. 31-5

FIBER-OPTIC RECEIVER, LOW SENSITIVITY, 2 μW

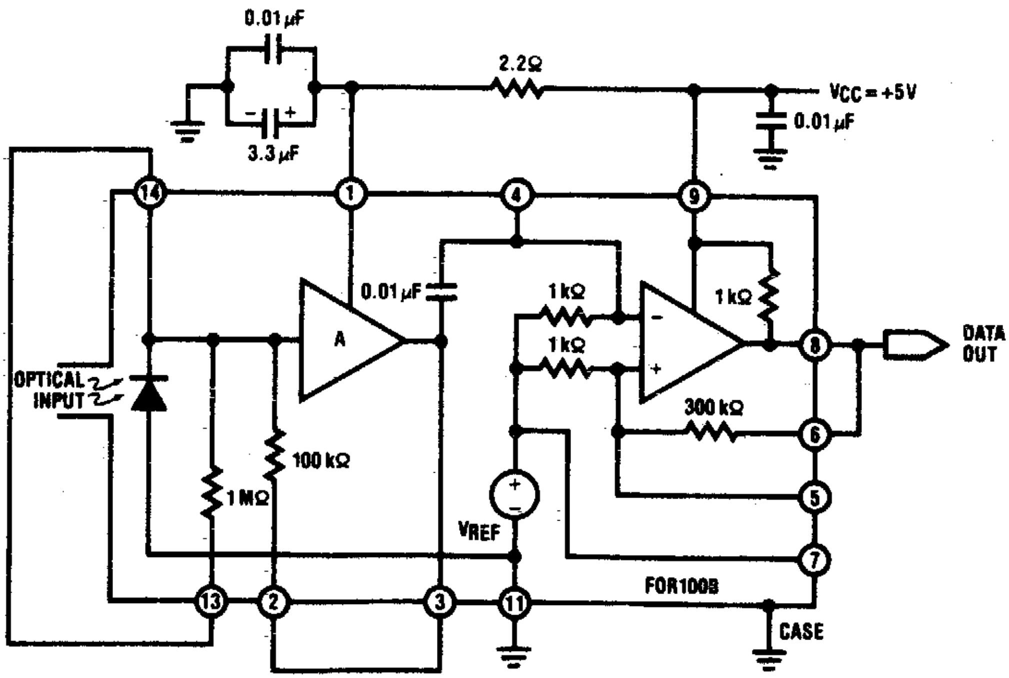


Fig. 31-6

32

Field Strength Meters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW COST MICROWAVE FIELD STRENGTH METER

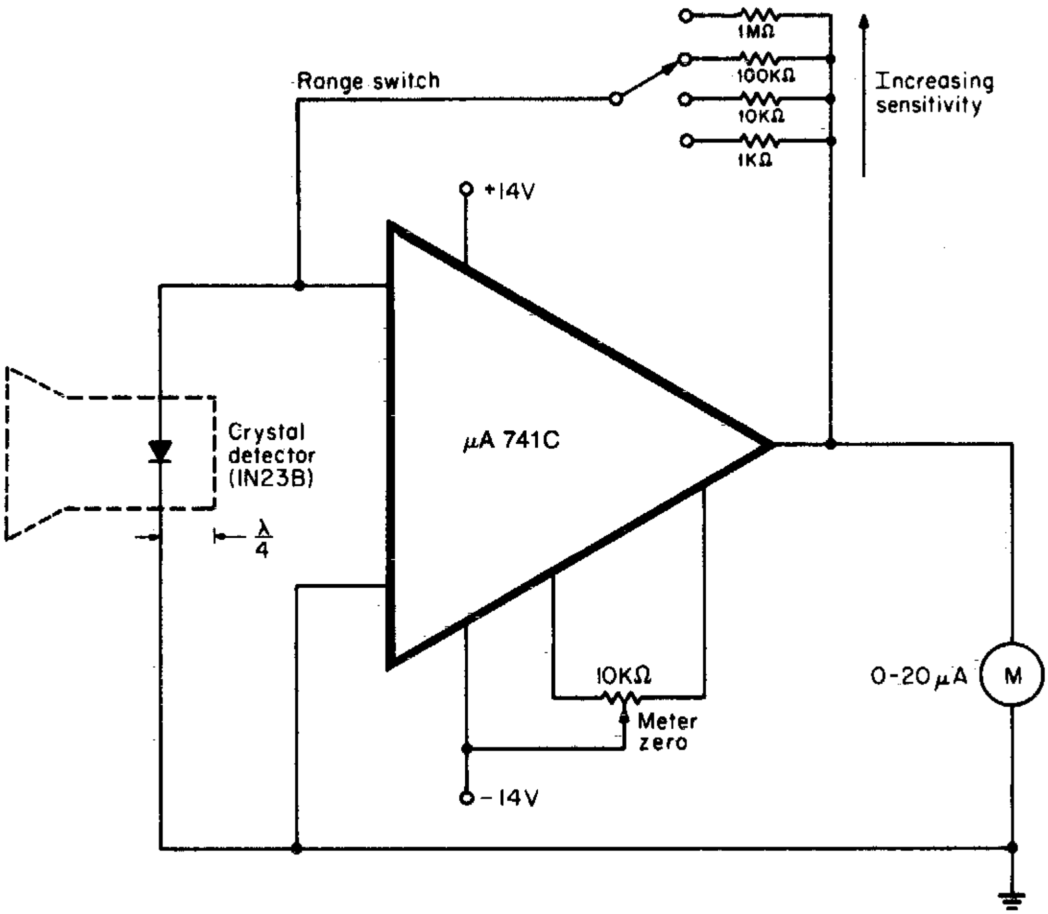


Fig. 32-1

Circuit Notes

When operating, a waveguide directs energy onto a crystal detector. The diode shown is for X-band operation. The waveguide is a 1½ inch piece of plastic tubing with the ends flared. The plastic is coated with an electroless copper solution to provide a conducting surface. The dimensions are not critical. For calibrated readings, the meter is placed in a known field or else compared to a calibrated meter. To operate the meter, point it away from the signal. Switch the meter to the desired range, and adjust the zero control for a 0 reading. Then point the waveguide at the signal, and read field strength directly.

SENSITIVE FIELD-STRENGTH METER

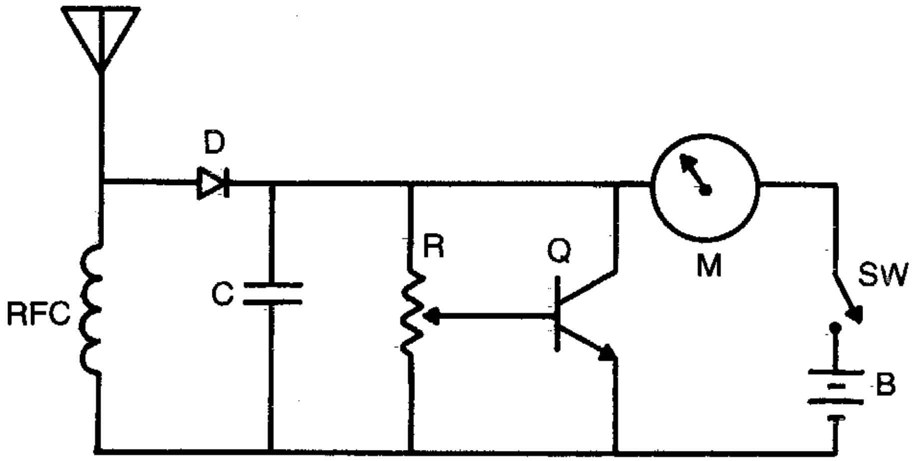
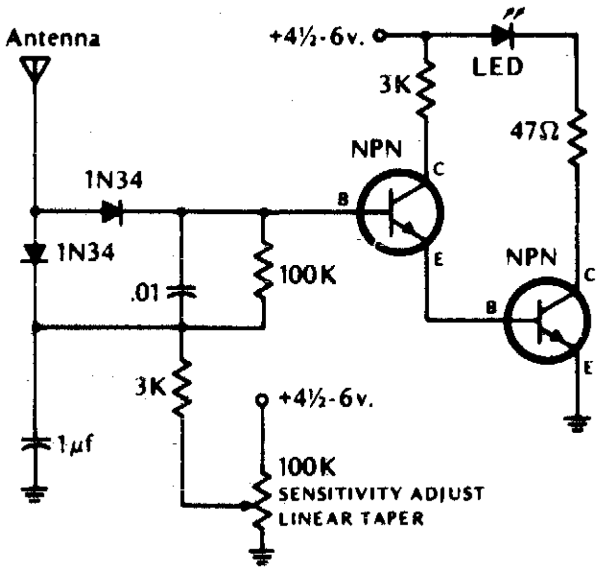


Fig. 32-2

Circuit Notes

Increased sensitivity gives field strength reading from low power transmitters. Operating range 3-30 MHz. To operate, adjust R for ⅓ to ½ scale reading. RFC = 2.5 mH choke, C = 1,000 pF, R = 50 K pot, M = 0 – 1 mA, D = 1N34 or 1N60 (Germanium), Q = NPN (RCASK3020, 2N3904 or equivalent).

ADJUSTABLE-SENSITIVITY FIELD-STRENGTH INDICATOR



Circuit Notes

The LED lights if the rf field is higher than the pre-set field strength level. Diodes should be germanium. Transistors (NPN) = 2N2222, 2N3393, 2N3904 or equivalent.

Fig. 32-3

FIELD STRENGTH METER – 1.5 to 150 MHz

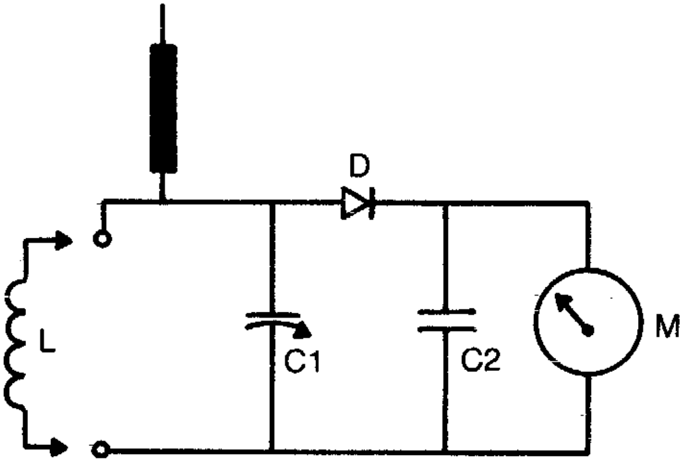
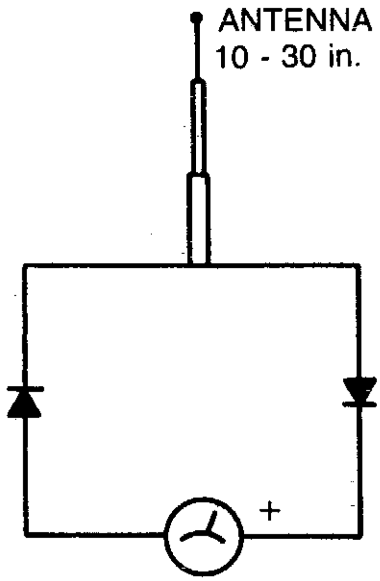


Fig. 32-4

Circuit Notes

The tuning range is determined by coil (L) dimensions and setting of C1. Coils can be plugged in for multirange use or soldered in place if only limited frequency range is of interest. C1 = 36 pF variable, C2 = .0047 disc, D = 1N60 (germanium) and M = 0–1 mA meter. For increased sensitivity, use 50 μA meter.

SIMPLE FIELD STRENGTH METER

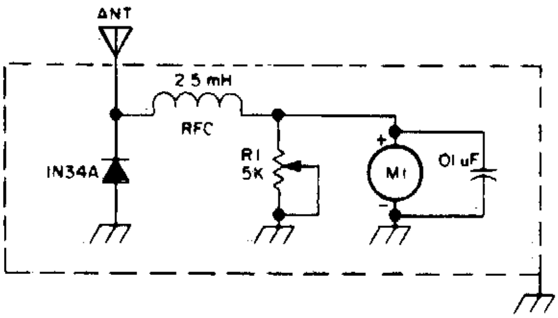


Circuit Notes

The circuit is frequency selective. It has been used from 2 meters through 160 meters. The telescoping antenna may be adjusted to its shortest length when working at 2 meters to keep the needle on the scale. Meter should be a 100 microamp to a 500 microamp movement. The diodes are germanium type, such as 1N34, etc. Silicon diodes will also work, but they are a bit less sensitive.

Fig. 32-5

UNTUNED FIELD STRENGTH METER

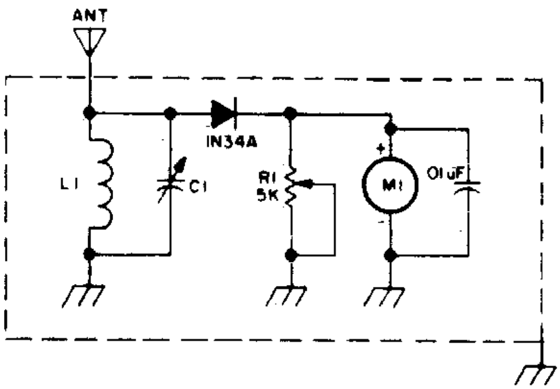


Circuit Notes

Sensitivity is controlled by R1 and sensitivity of Meter M1.

Fig. 32-6

TUNED FIELD STRENGTH METER



Circuit Notes

Resonant combination of L1 and C1 are selected to cover frequencies desired.

Fig. 32-7

VOM FIELD STRENGTH METER

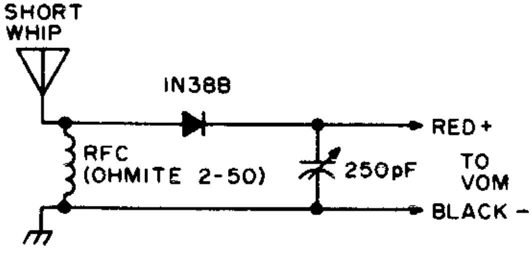


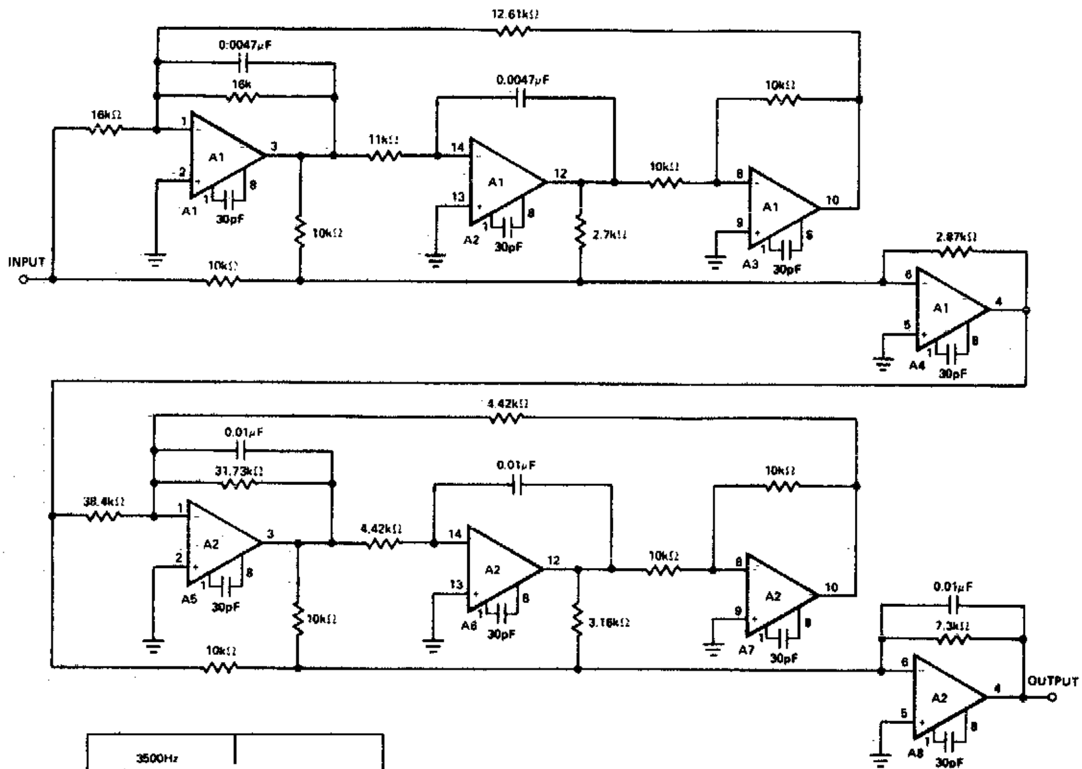
Fig. 32-8

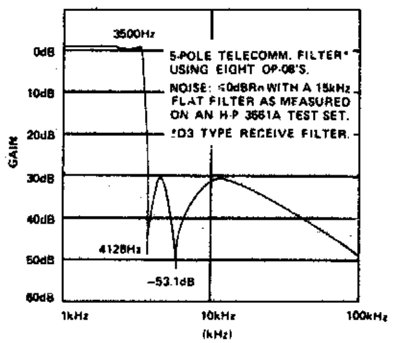
33

Filters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FIVE-POLE ACTIVE FILTER





The above realization of a type D3 receive filter is accomplished using eight OP-08's. As can be seen from the response curve, the >30dB attenuation in the stop band requirement has been met. In addition, the noise performance of <0dBRn has been measured. One of the unique features of the OP-08 is its low supply current of 600μA maximum. Thus the total supply drain for all eight op amps is only 4.8mA.

Fig. 33-1

DIGITALLY TUNED LOW POWER ACTIVE FILTER

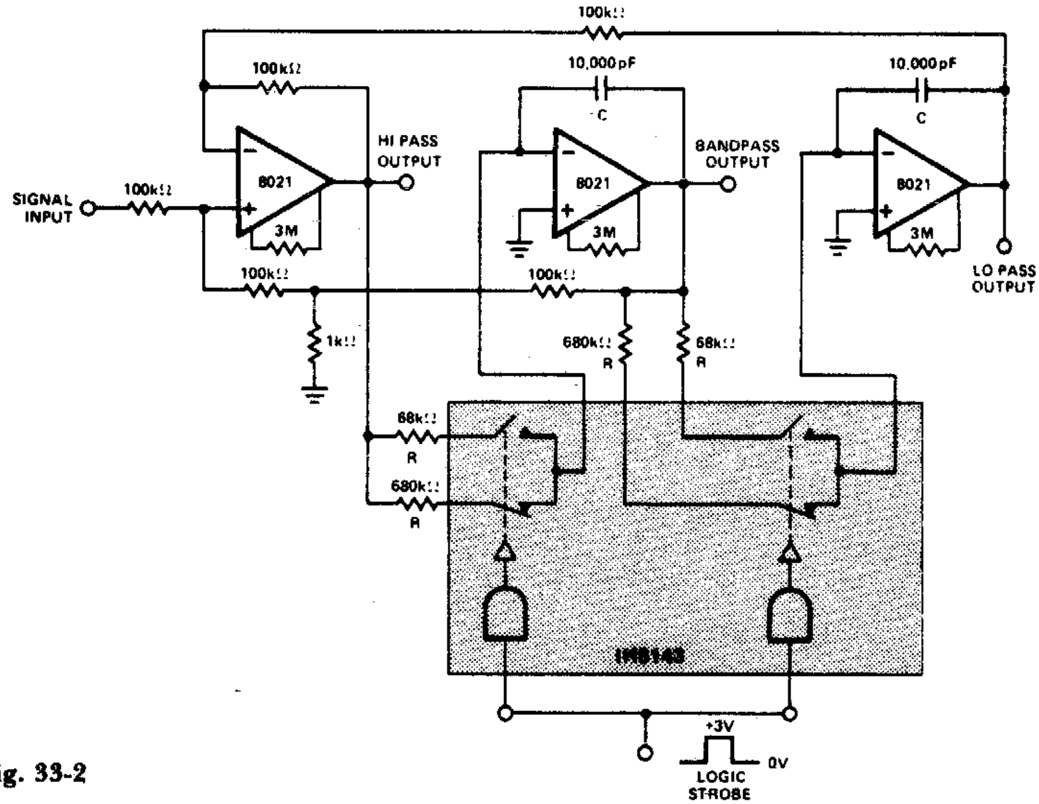


Fig. 33-2

Circuit Notes

Constant gain, constant Q, variable frequency filter which provides simultaneous low-pass, bandpass, and high-pass outputs. With the component values shown, center frequency will be 235 Hz and 23.5 Hz for high and low logic inputs respectively, Q = 100, and gain = 100.

$$f_n = \text{center frequency} = \frac{1}{2\pi RC}$$

10 kHz SALLEN-KEY LOW-PASS FILTER

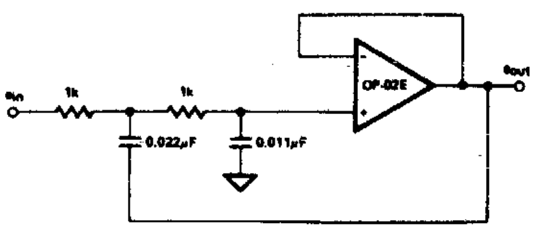
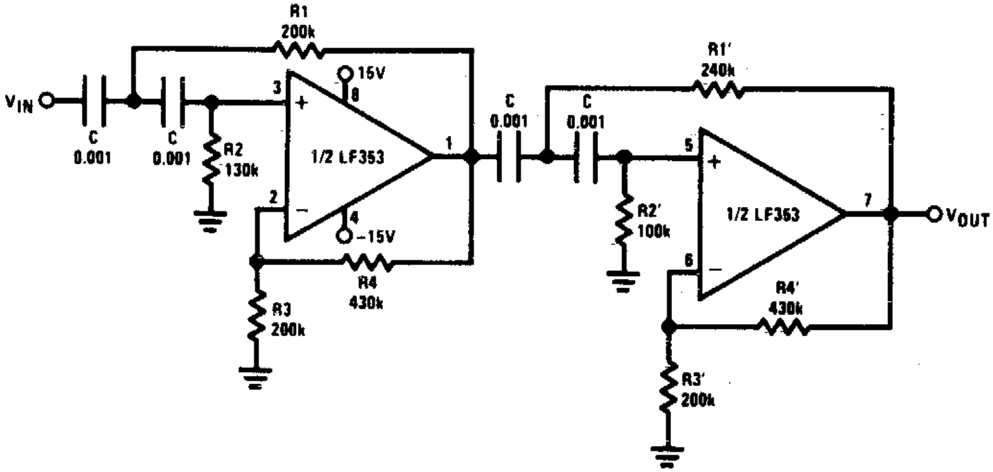


Fig. 33-3

FOURTH ORDER HIGH-PASS BUTTERWORTH FILTER



- Corner frequency (f_c) = $\sqrt{\dfrac{1}{R1R2C^2}} \cdot \dfrac{1}{2\pi} = \sqrt{\dfrac{1}{R1'R2'C^2}} \cdot \dfrac{1}{2\pi}$
- Passband gain (H_O) = (1 + R4/R3)(1 + R4'/R3')
- First stage Q = 1.31
- Second stage Q = 0.541
- Circuit shown uses closest 5% tolerance resistor values for a filter with a corner frequency of 1 kHz and a passband gain of 10

Fig. 33-4

TUNABLE NOTCH FILTER TO SUPPRESS HUM

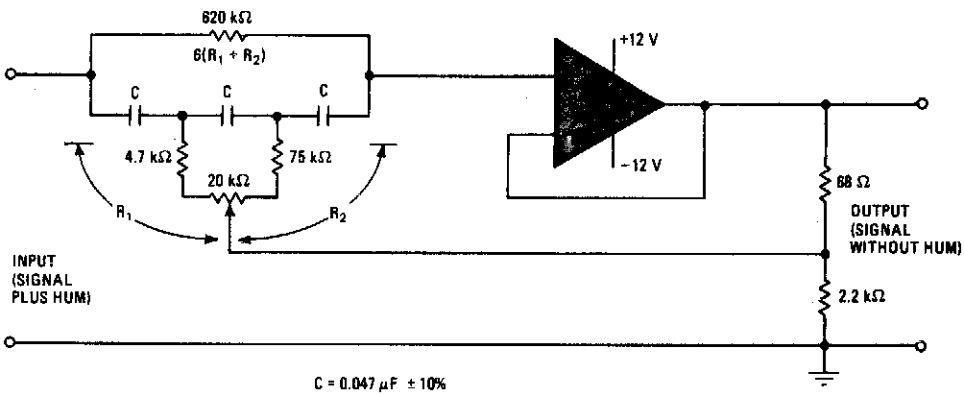
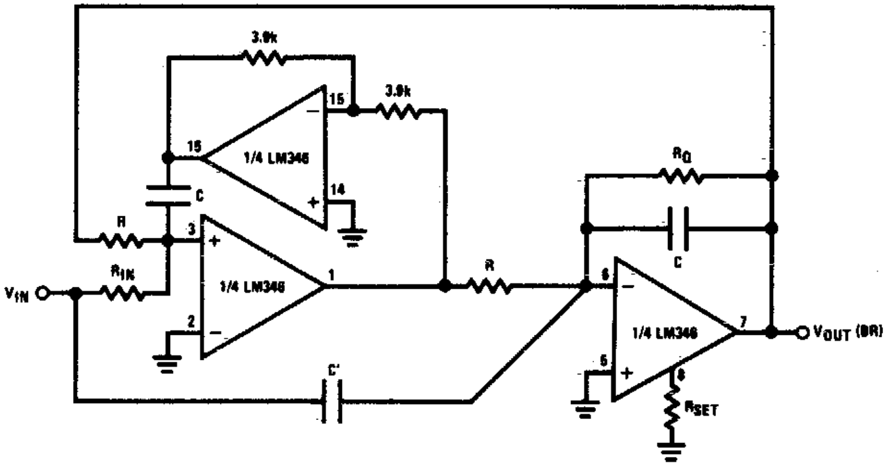


Fig. 33-5

Circuit Notes

This narrow-stop-band filter can be tuned by the pot to place the notch at any frequency from 45 to 90 Hz. It attenuates power-line hum or other unwanted signals by at least 30 dB. Because the circuit uses wide-tolerance parts, it is inexpensive to build.

THREE-AMPLIFIER NOTCH FILTER (OR ELLIPTIC FILTER BUILDING BLOCK)



Circuit Synthesis Equations

$$R \times C = \frac{0.159}{f_o}\ ;\ R_Q = Q_o \times R;\ R_{IN} = \frac{0.159 \times f_o}{C' \times f^2_{notch}}$$

- For nothing but a notch output: $R_{IN} = R$, $C' = C$.

$$H_{o(BR)}\Big|_{f << f_{notch}} = \frac{R}{R_{IN}}\quad H_{o(BR)}\Big|_{f >> f_{notch}} = \frac{C'}{C}$$

Fig. 33-6

SELECTABLE BANDWIDTH NOTCH FILTER

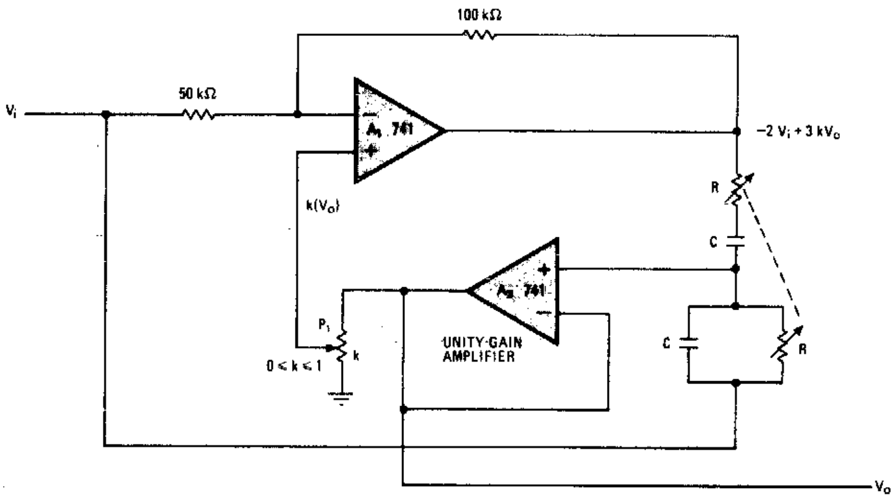


Fig. 33-7

Circuit Notes

This notch filter, which operates at up to 200 kHz, uses a modified Wien bridge to select bandwidth over which frequencies are rejected. RC components determine filter's center frequency, P1 selects notch bandwidth. Notch depth is fixed at about 60 dB.

4.5 MHz NOTCH FILTER

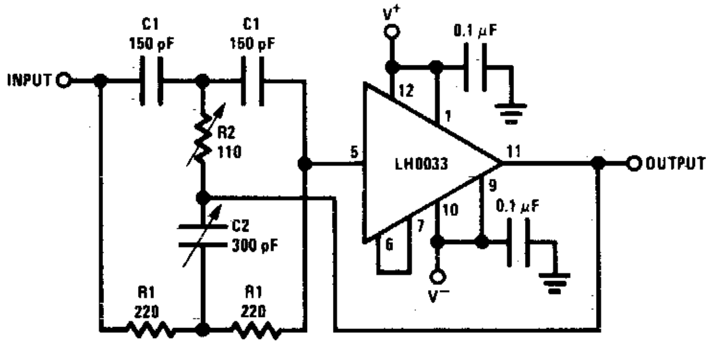
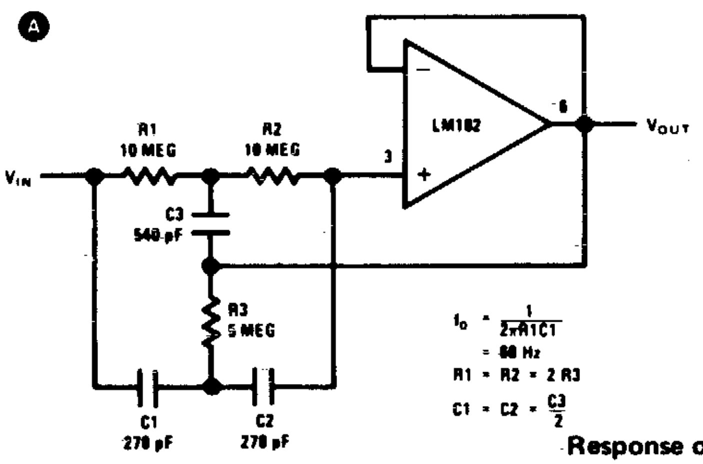


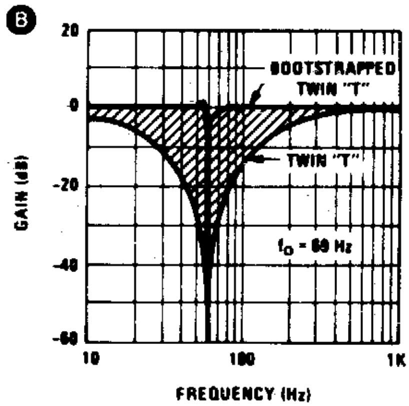
Fig. 33-8

Circuit Notes

Component value sensitivity is extremely critical, as are temperature coefficients and matching of the components. Best performance is attained when perfectly matched components are used and when the gain of the amplifier is unity. To illustrate, the quality factor Q is very high as amplifier gain approaches 1 with all components matched (in fact, theoretically it approaches ∞) but decreases to about 12.5 with the amplifier gain at 0.98.

HIGH Q NOTCH FILTER





Response of High and Low Q Notch Filter

Fig. 33-9

Circuit Notes

A shows a twin-T network connected to an LM102 to form a high Q, 60 Hz notch filter. The junction of R3 and C3, which is normally connected to ground, is bootstrapped to the output of the follower. Because the output of the follower is a very low impedance, neither the depth nor the frequency of the notch change; however, the Q is raised in proportion to the amount of signal fed back to R3 and C3. B shows the response of a normal twin-T and the response with the follower added.

REJECTION FILTER

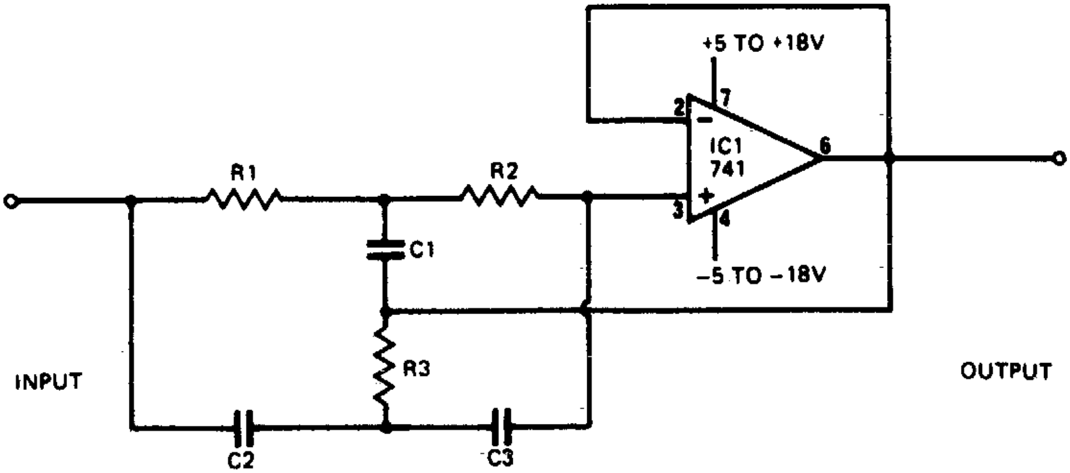


Fig. 33-10

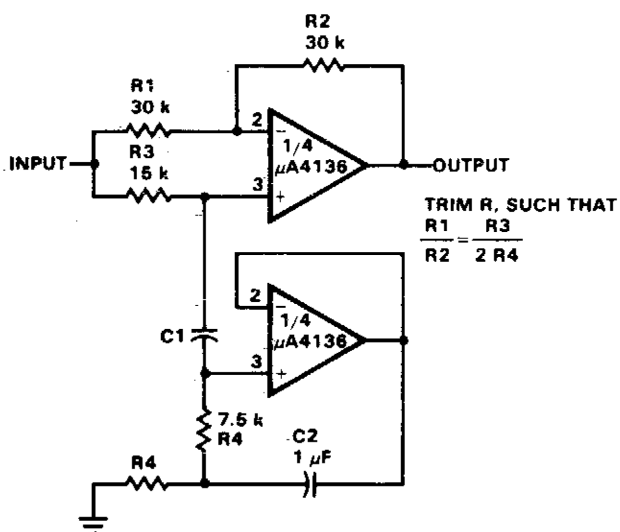
Circuit Notes

This narrowband filter using the 741 operational amplifier can provide up to 60 dB of rejection. With resistors equal to 100 K and capacitors equal to 320 pF, the circuit will reject 50 Hz. Frequencies within the range 1 Hz to 10 kHz may be rejected by selecting components in accordance with the formula:

$$F = \frac{1}{2\pi RC}$$

To obtain rejections better than 40 dB, resistors should be matched to 0.1% and capacitors to 1%.

NOTCH FILTER USING THE μA4136 AS A GYRATOR



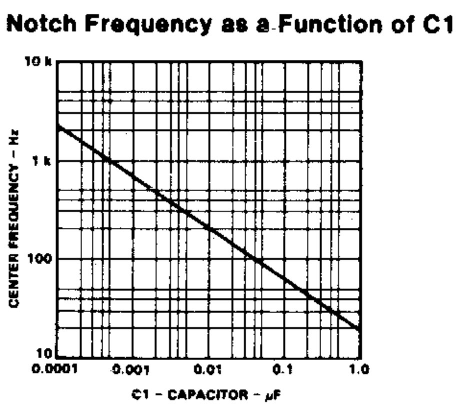


Fig. 33-11

1 kHz BANDPASS ACTIVE FILTER

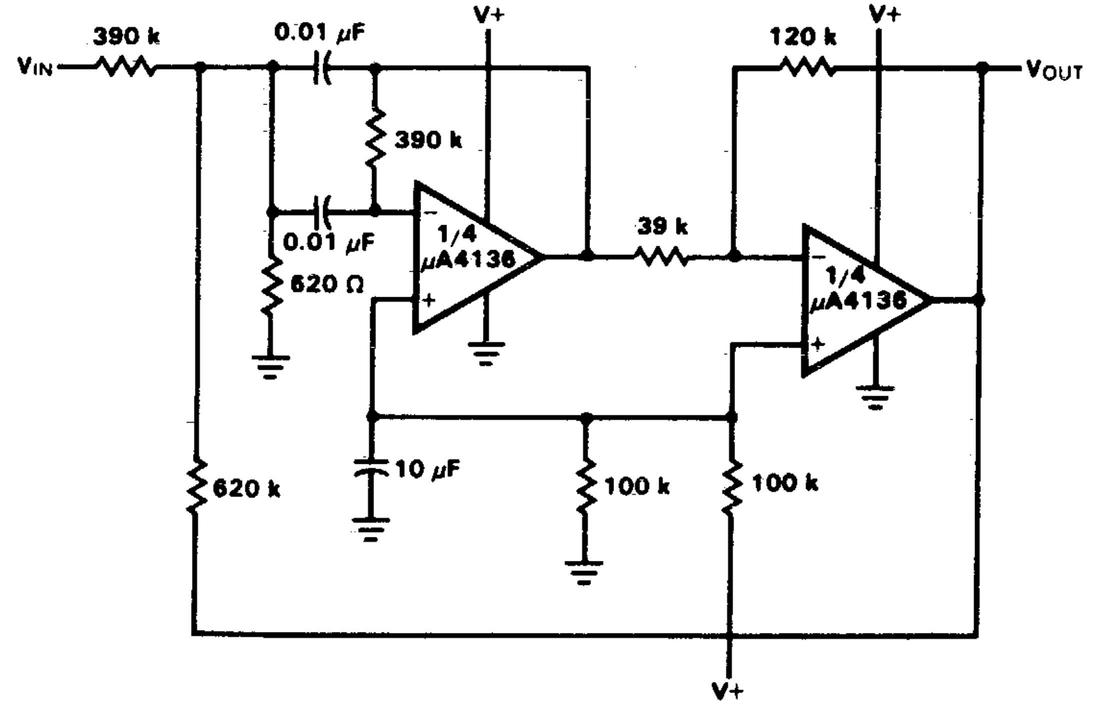
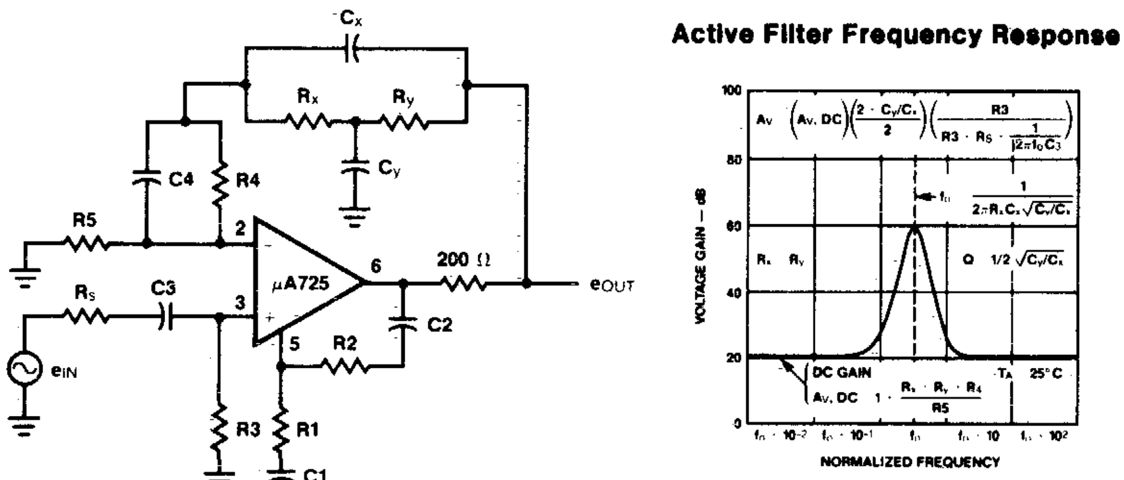


Fig. 33-12

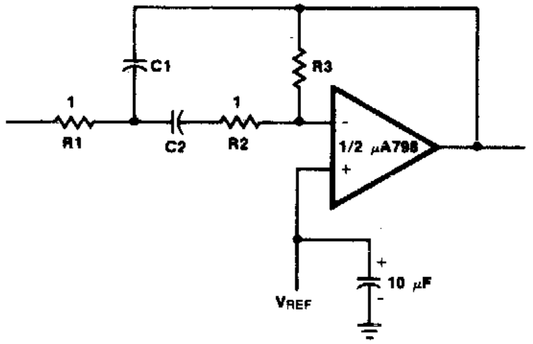
BANDPASS ACTIVE FILTER WITH 60 dB GAIN



Pin numbers are shown for metal package only.

Fig. 33-13

MULTIPLE FEEDBACK BANDPASS FILTER



$f_o \stackrel{\Delta}{=}$ center frequency

$BW \stackrel{\Delta}{=}$ Bandwidth

R in kΩ

C in μF

$Q = \frac{f_o}{BW} < 10$

$C1 = C2 = \frac{Q}{3}$

$R1 = R2 = 1$
$R3 = 9Q^2 - 1$ } Use scaling factors in these expressions

If source impedance is high or varies, filter may be preceded with voltage follower buffer to stabilize filter parameters.

Design example:

given: $Q = 5$, $f_o = 1$ kHz

Let $R1 = R2 = 10$ kΩ

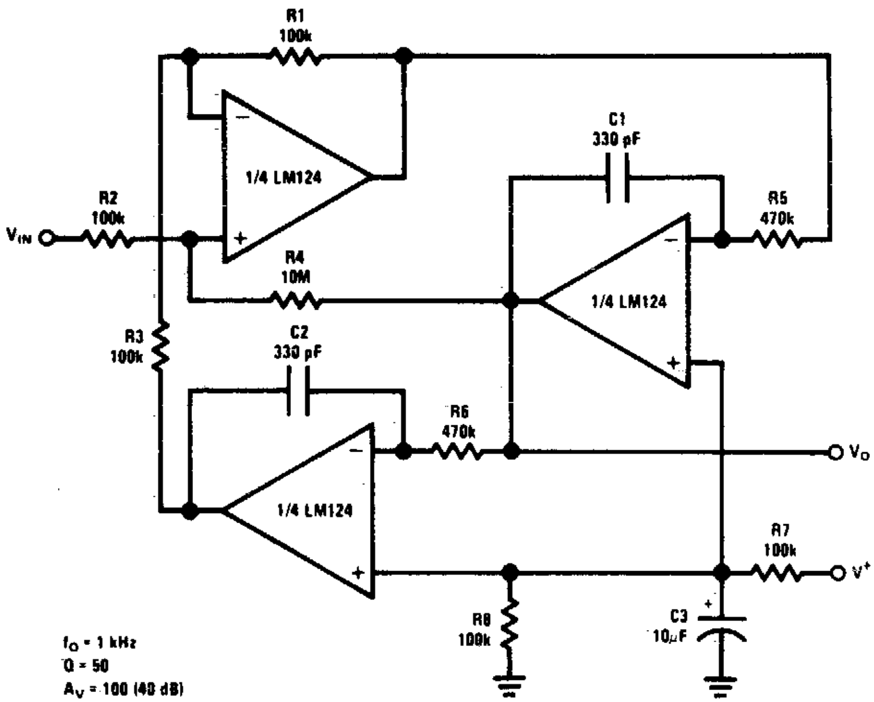
then $R3 = 9(5)^2 - 10$

$R3 = 215$ kΩ

$C = \frac{5}{3} = 1.6$ nF

Fig. 33-14

BIQUAD RC ACTIVE BANDPASS FILTER



$f_O = 1$ kHz
$Q = 50$
$A_V = 100$ (40 dB)

Fig. 33-15

400 Hz LOW-PASS BUTTERWORTH ACTIVE FILTER

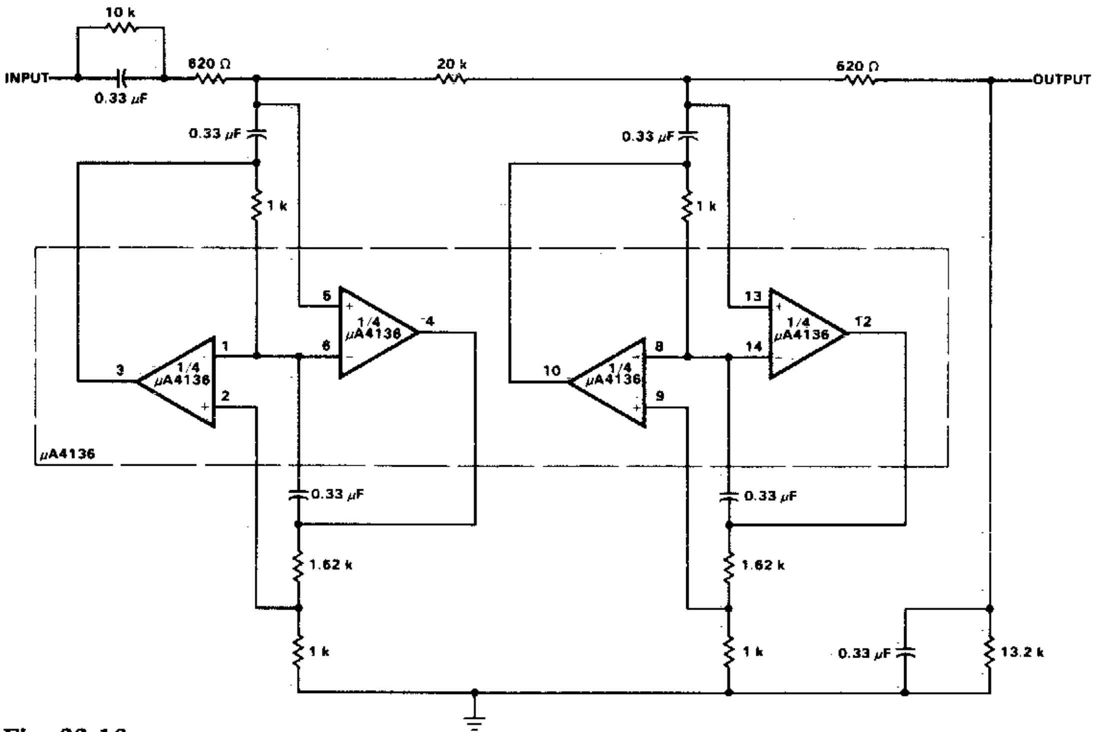


Fig. 33-16

VARIABLE BANDWIDTH BANDPASS ACTIVE FILTER

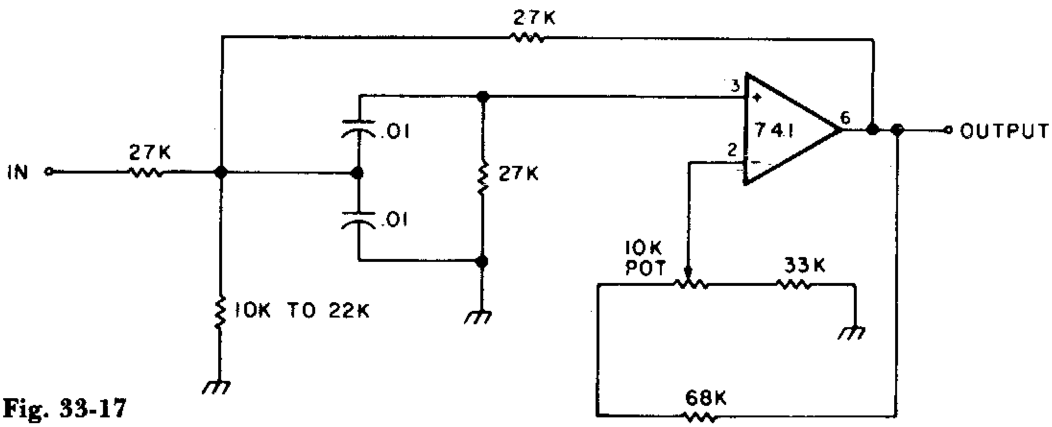


Fig. 33-17

Circuit Notes

This circuit has adjustable bandwidth with values for a center frequency of about 800 Hz. The 10 K pot adjusts bandwidth from approximately ±350 Hz to ±140 Hz at 3 dB down points.

LOW-PASS FILTER

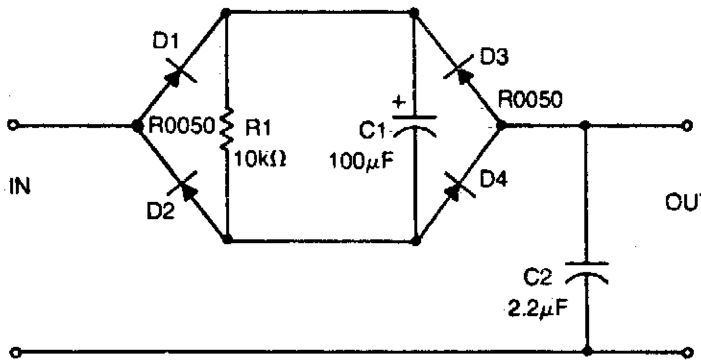


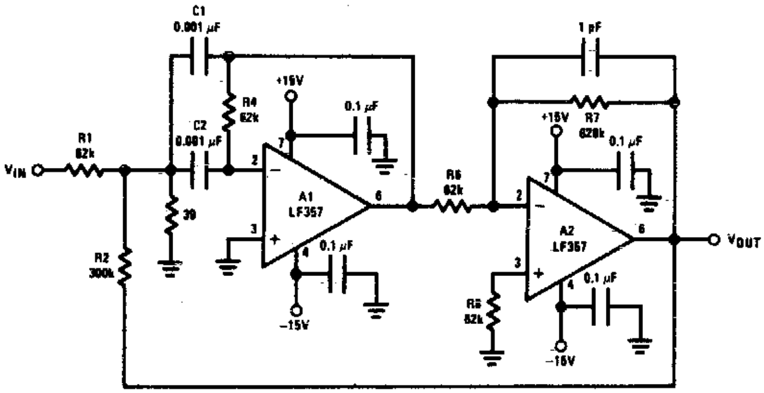
Fig. 33-18

D1, D2, D3, D4—HEP R0050 C2—2.2μF
C1—100μF, 50V electrolytic R1—10kΩ, 1/2W

Circuit Notes

This nonlinear, passive filter circuit rejects ripple (or unwanted but fairly steady voltage) without appreciably affecting the rise time of a signal. The circuit works best when the signal level is considerably lower than the unwanted ripple, provided the ripple level is fairly constant. The circuit has characteristics similar to two peak-detecting sample-and-hold circuits in tandem with a voltage averager.

HIGH Q BANDPASS FILTER



- By adding positive feedback (R2) Q increases to 40
- f_{BP} = 100 kHz

$$\frac{V_{OUT}}{V_{IN}} = 10\sqrt{Q}$$

- Clean layout recommended
- Response to a 1 Vp-p tone burst: 300 μs

Fig. 33-19

MFB BANDPASS FILTER FOR MULTICHANNEL TONE DECODER

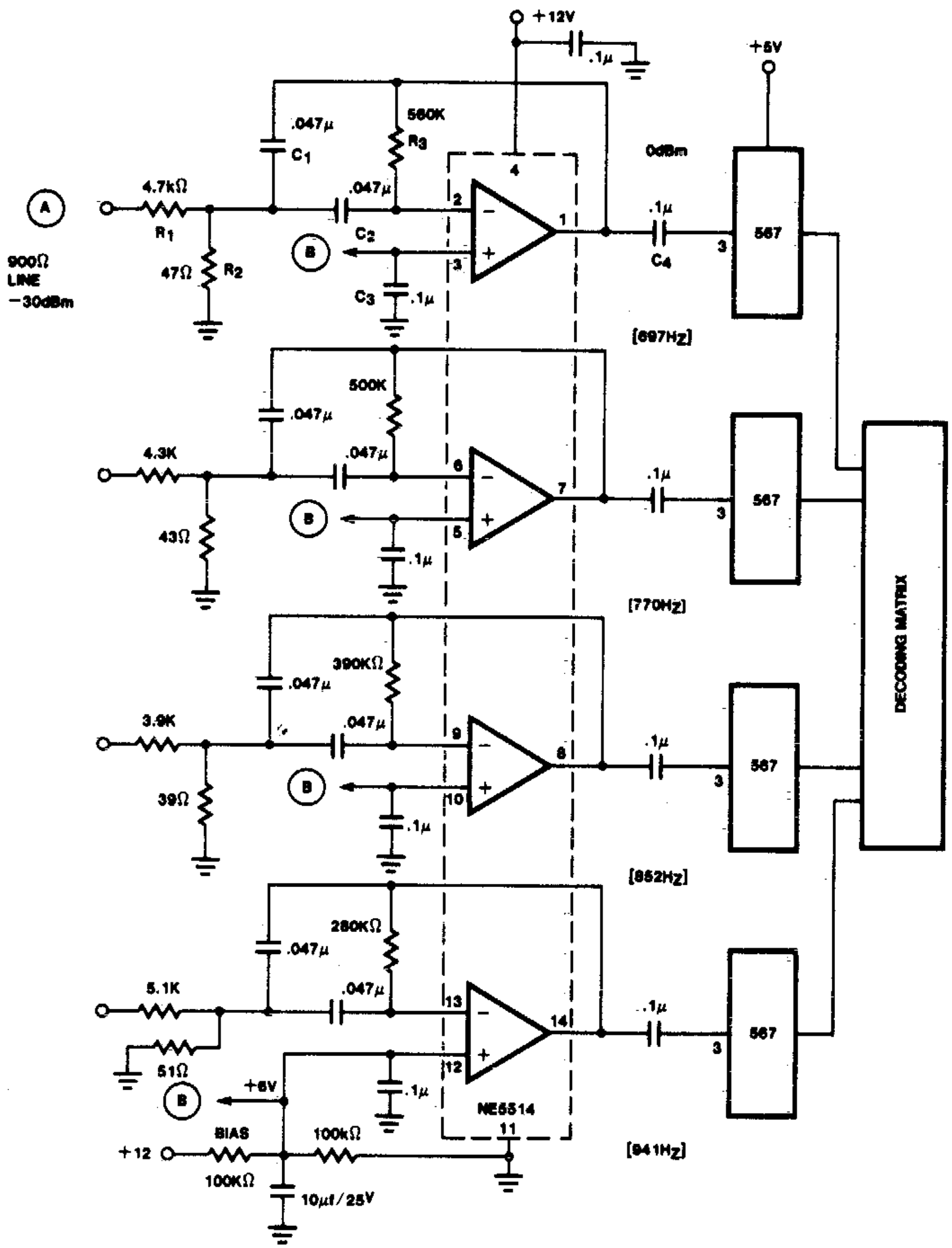
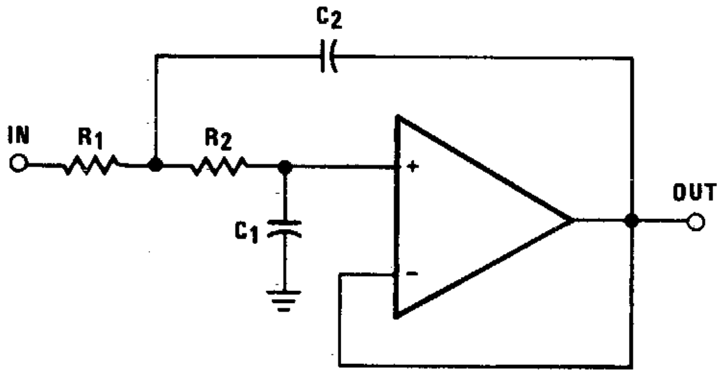


Fig. 33-20

SALLEN-KEY SECOND ORDER LOW-PASS FILTER

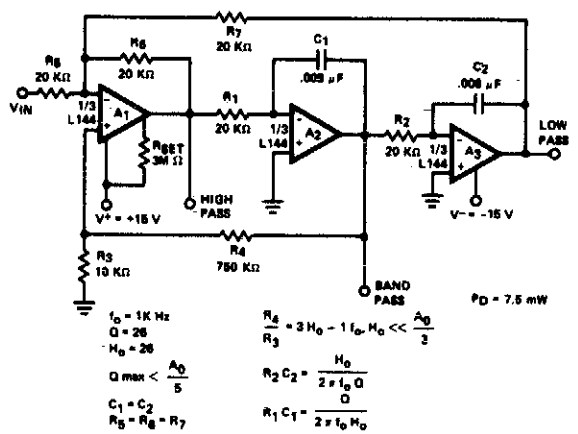


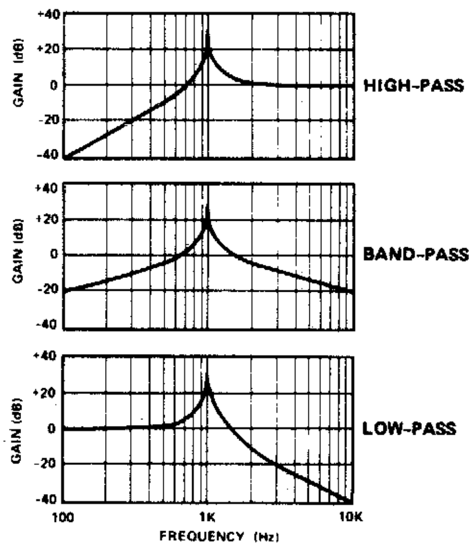
NOTES:

1. Make $R_1 = R_2$
2. $fc = \frac{1}{2\ R_1\sqrt{C_1C_2}}$
3. $Q = \frac{1}{2}\sqrt{\frac{C_2}{C_1}}$

Fig. 33-21

THREE AMPLIFIER ACTIVE FILTER





Bode plots of Active Filter Output

Fig. 33-22

Circuit Notes

The active filter is a state variable filter with bandpass, high-pass and low-pass outputs. It is a classical analog computer method of implementing a filter using three amplifiers and only two capacitors.

BANDPASS STATE VARIABLE FILTER

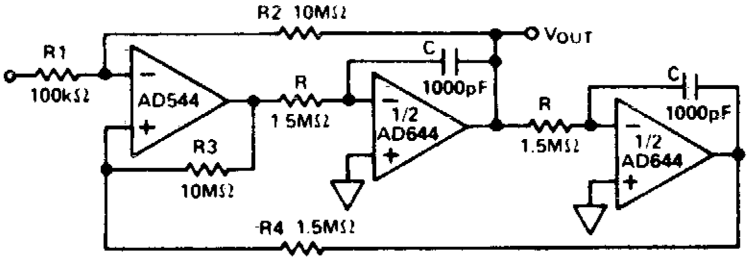


Fig. 33-23

f_o = CENTER FREQUENCY = $1/2\pi RC$

Q_o = QUALITY FACTOR = $\frac{R_1 + R_2}{2R_1}$

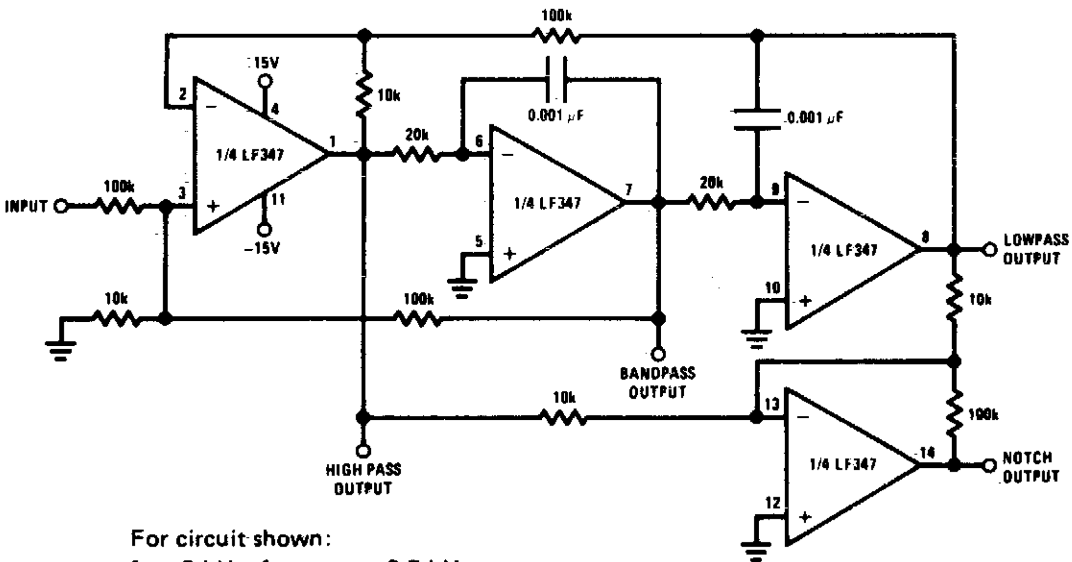
H_o = GAIN AT RESONANCE = R_2/R_1

$R_3 = R_4 \approx 10^8/f_o$

Qo, IS ADJUSTABLE BY VARYING R2

fo, IS ADJUSTABLE BY VARYING R OR C

UNIVERSAL STATE VARIABLE FILTER



For circuit shown:

f_O = 3 kHz, f_{NOTCH} = 9.5 kHz

Q = 3.4

Passband gain:

- Highpass — 0.1
- Bandpass — 1
- Lowpass — 1
- Notch — 10

- $f_O \times Q \leq 200$ kHz
- 10V peak sinusoidal output swing without slew limiting to 200 kHz
- See LM348 data sheet for design equations

Fig. 33-24

500 Hz SALLEN-KEY BANDPASS FILTER

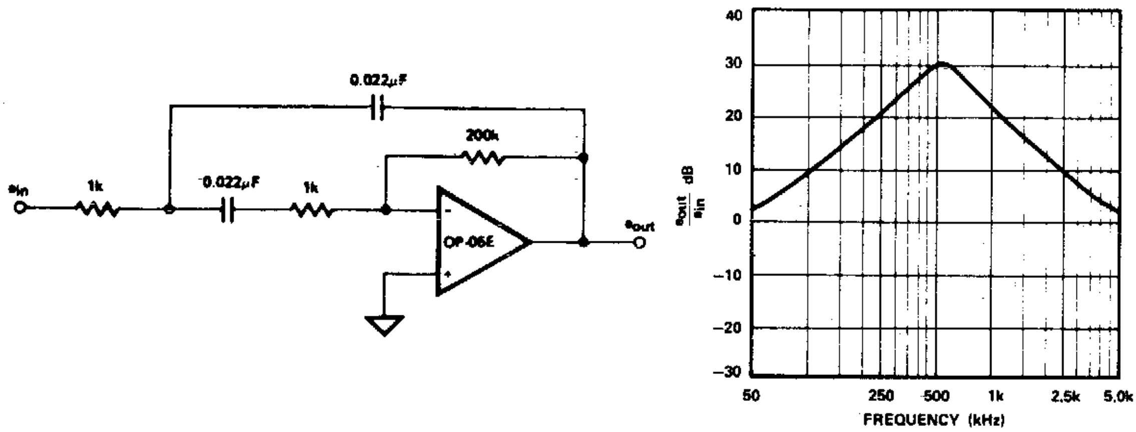
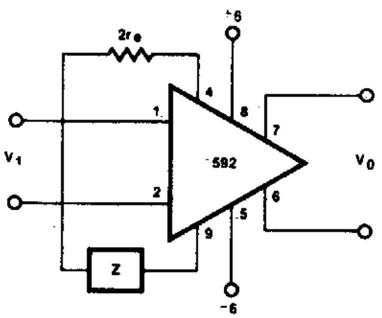


Fig. 33-25

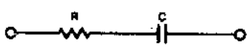
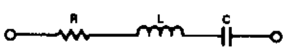
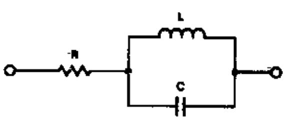
FILTER NETWORKS



$$\frac{V_0(s)}{V_1(s)} \cong \frac{1.4 \times 10^4}{Z(s) + 2r_e}$$

$$\cong \frac{1.4 \times 10^4}{Z(s) + 32}$$

BASIC CONFIGURATION

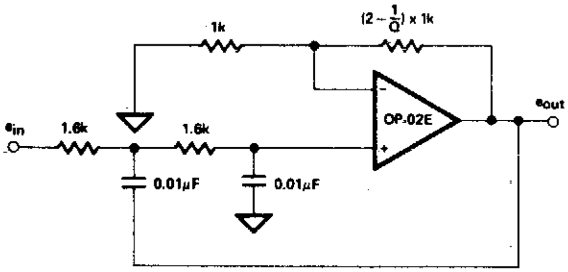
Z NETWORK	FILTER TYPE	$\frac{V_0(s)}{V_1(s)}$ TRANSFER FUNCTION
	LOW PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{1}{s + R/L}\right]$
	HIGH PASS	$\frac{1.4 \times 10^4}{R}\left[\frac{s}{s + 1/RC}\right]$
	BAND PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{s}{s^2 + R/L\, s + 1/LC}\right]$
	BAND REJECT	$\frac{1.4 \times 10^4}{R}\left[\frac{s^2 + 1/LC}{s^2 + 1/LC + s/RC}\right]$

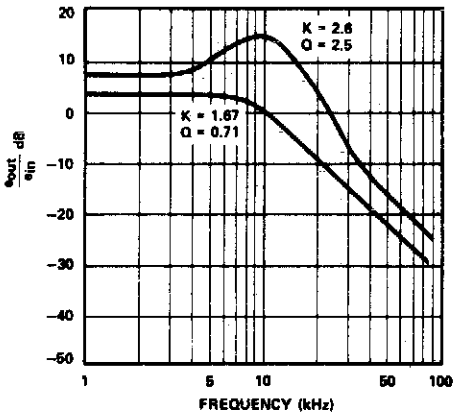
NOTE

In the networks above, the R value used is assumed to include $2r_e$, or approximately 32Ω.

Fig. 33-26

EQUAL COMPONENT SALLEN-KEY LOW-PASS FILTER

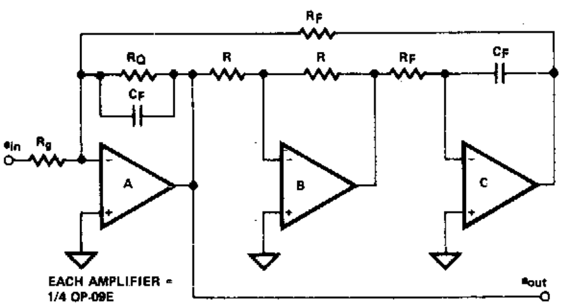




Equal R, Equal C Sallen-Key Response

Fig. 33-27

BIQUAD FILTER



Circuit Notes

The biquad filter, while appearing very similar to the state-variable filter, has a bandwidth that is fixed regardless of center frequency. This type of filter is useful in applications such as spectrum analyzers, which require a filter with a fixed bandwidth.

Fig. 33-28

SECOND ORDER STATE VARIABLE FILTER (1 kHz, Q = 10)

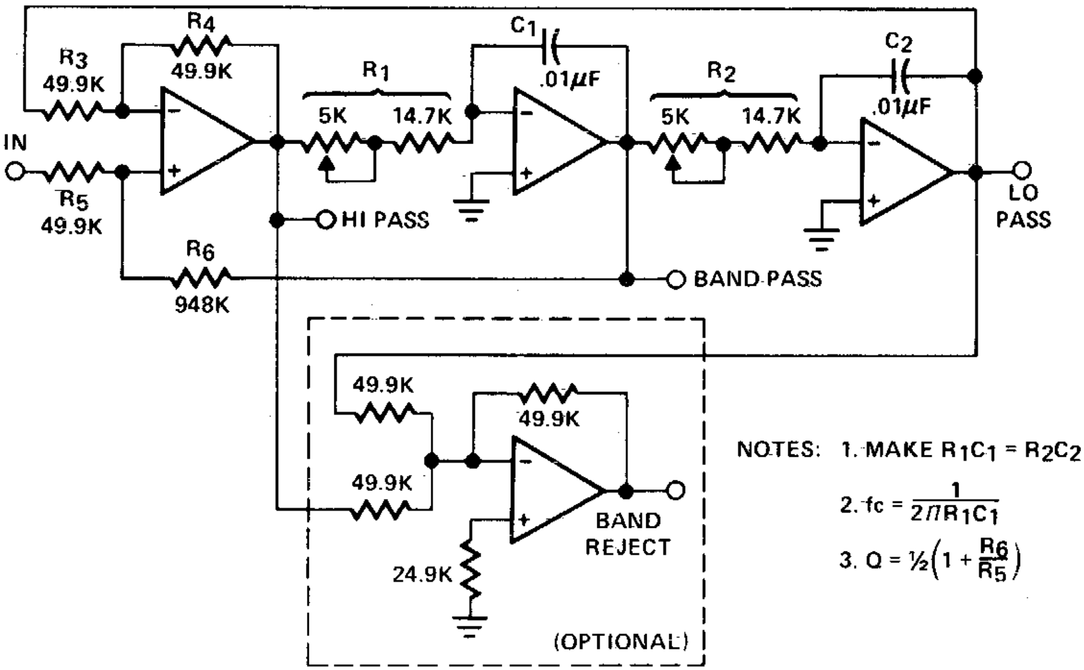


Fig. 33-29

BIQUAD FILTER

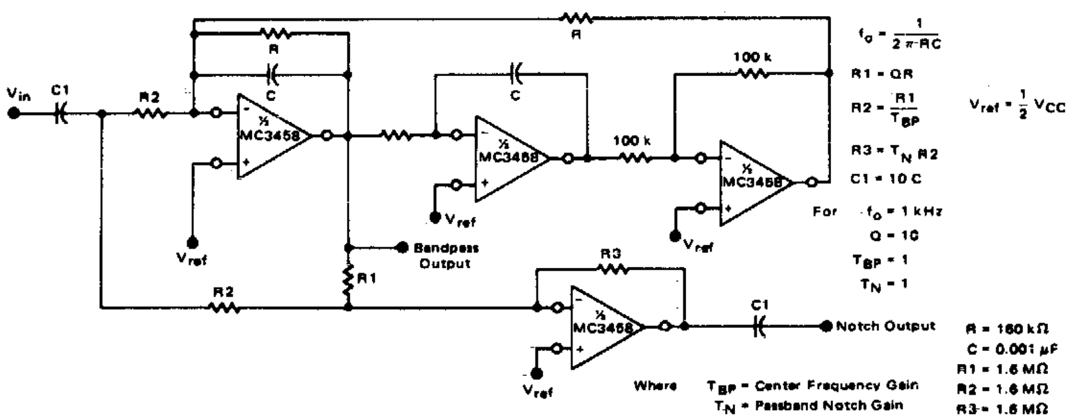


Fig. 33-30

TUNABLE ACTIVE FILTER

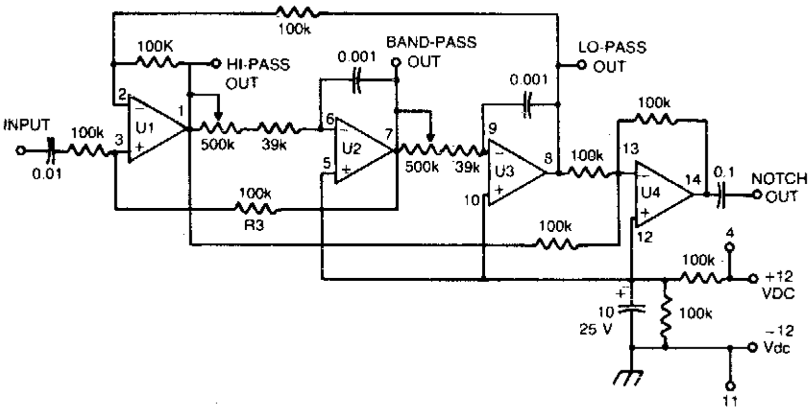
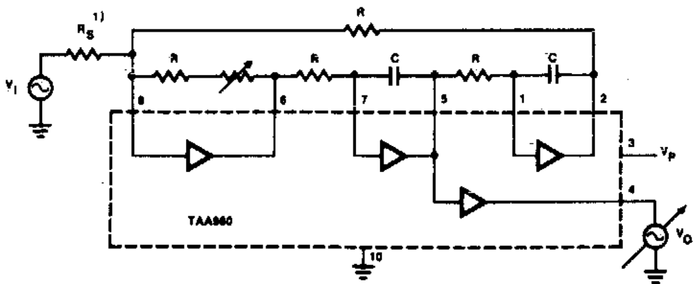


Fig. 33-31

Circuit Notes

The high-pass and low-pass outputs covering the range of 300 Hz to 3000 Hz have been summed in the fourth op amp to provide a notch output. The potentiometers must have a reverse log taper. Fixed-frequency active filter center frequency is 1 kHz, with a Q of 50.

ACTIVE RC FILTER FOR FREQUENCIES UP TO 150 kHz



R = 10kΩ

This frequency range can be extended to 200kHz if a feed forward capacitor is connected between pin 5 and 8.

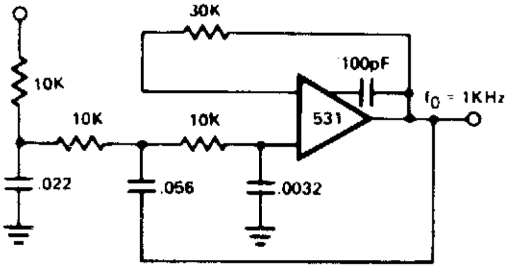
Fig. 33-32

f	Frequency	$\frac{1}{2\pi RC}$	
V_P	Supply voltage	6	V
	Filter performance		
Q	at T_A = 25°C	40 to 55	
Q	at T_A = −30 to +65°C	35 to 55	
V_i	Input voltage	400	mV
V_o	Output voltage	400	mV
d_{tot}	Distortion at V_o = 350mV	2	%
S/N	S/N ratio at V_o = 400mV	50	dB
Rs	Input resistor*	470	kΩ

*NOTE

Value of input resistor to be determined for $\frac{V_o}{V_i}$ = 0.90 to 1.1.

POLE ACTIVE LOW-PASS FILTER (BUTTERWORTH MAXIMALLY FLAT RESPONSE)



*Reference—EDN Dec. 15, 1970
Simplify 3-Pole Active Filter Design
A. Paul Brokow

RESPONSE OF 3-POLE ACTIVE BUTTERWORTH MAXIMALLY FLAT FILTER

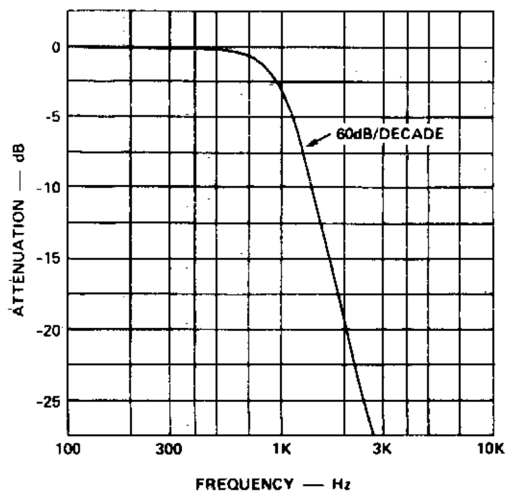
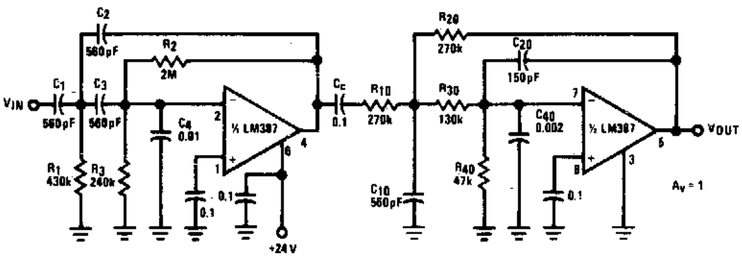
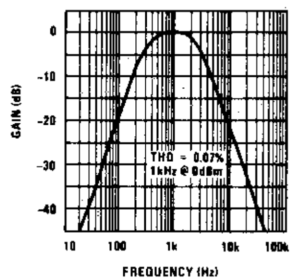


Fig. 33-33

SPEECH FILTER (300 Hz .3 kHz BANDPASS)





Speech Filter Frequency Response

Fig. 33-34

0.1 Hz TO 10 Hz BANDPASS FILTER

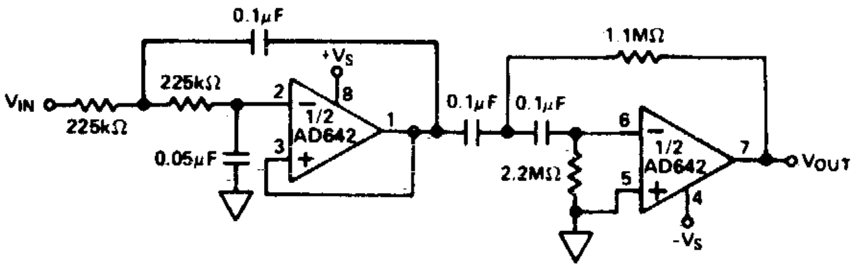


Fig. 33-35

HIGH-PASS ACTIVE FILTER

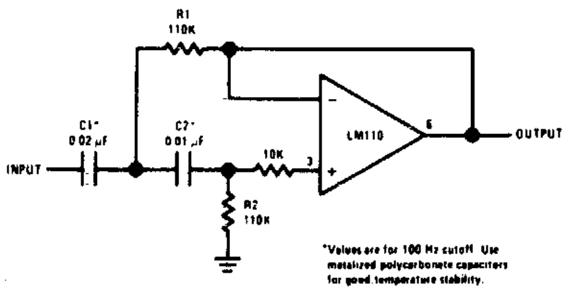
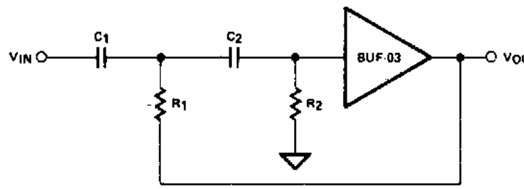


Fig. 33-36

HIGH-PASS FILTER (HIGH FREQUENCY)



$$\omega_o = \left(\frac{1}{R_1R_2C_1C_2}\right)^{1/2}$$

IF $C_1 = C_2 = C$, THEN

$$Q = \frac{(R_1/R_2)^{1/2}}{2}$$

f_o Hz	C pF	R_1 Ω	R_2 Ω	Q
500K	220	2.05K	1.02K	0.71

Fig. 33-38

SECOND ORDER HIGH-PASS ACTIVE FILTER

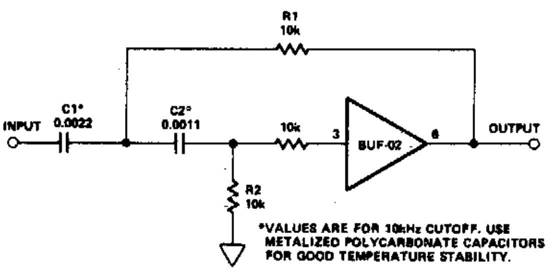


Fig. 33-37

160 Hz BANDPASS FILTER

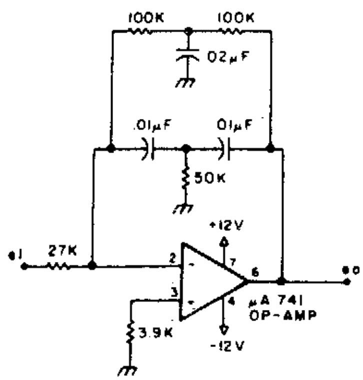


Fig. 33-39

MULTIPLE FEEDBACK BANDPASS FILTER (1.0 kHz)

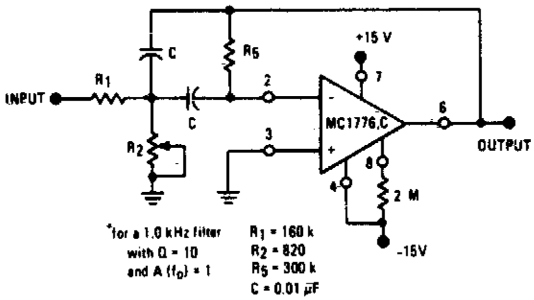


Fig. 33-40

RUMBLE FILTER USING LM387

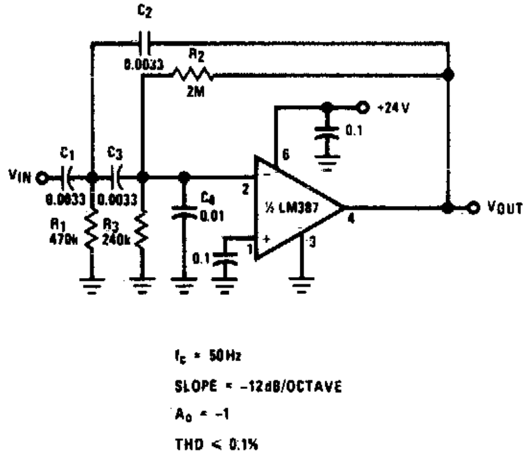


Fig. 33-42

20 kHz BANDPASS ACTIVE FILTER

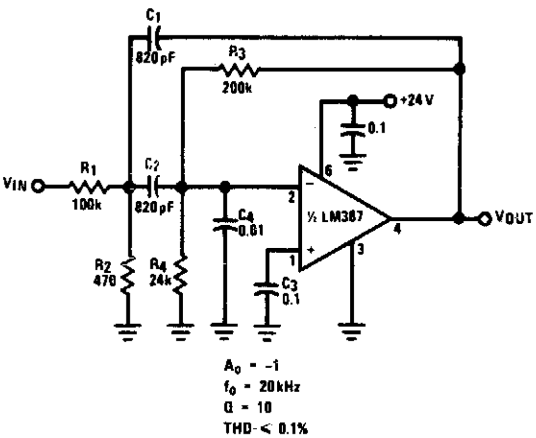


Fig. 33-41

SCRATCH FILTER USING LM387

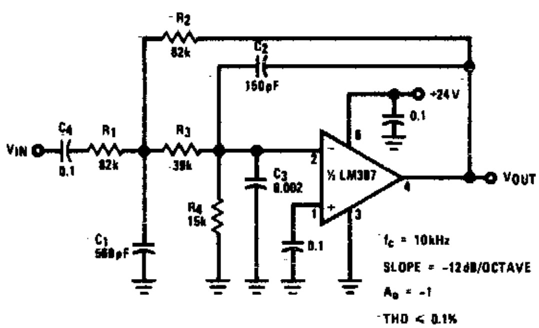


Fig. 33-43

34

Flashers and Blinkers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AUTO, BOAT, OR BARRICADE FLASHER

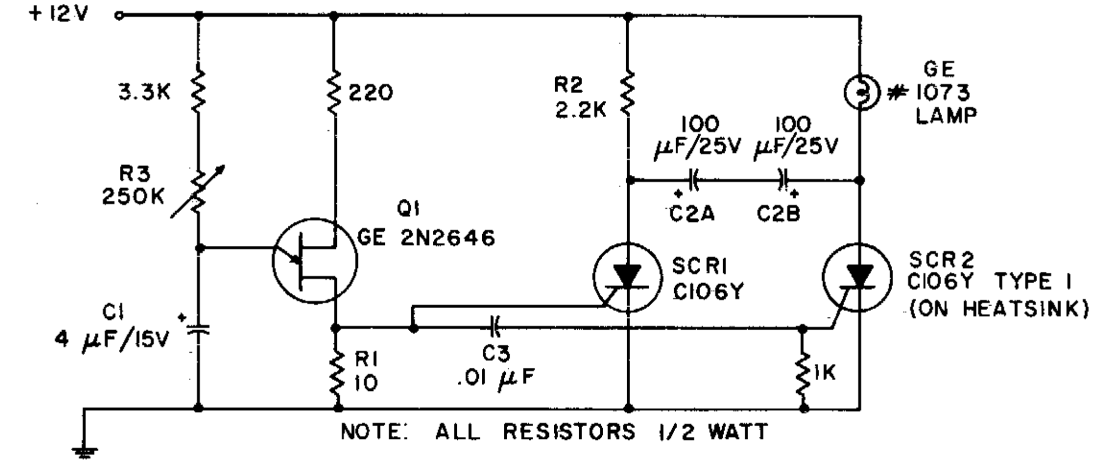
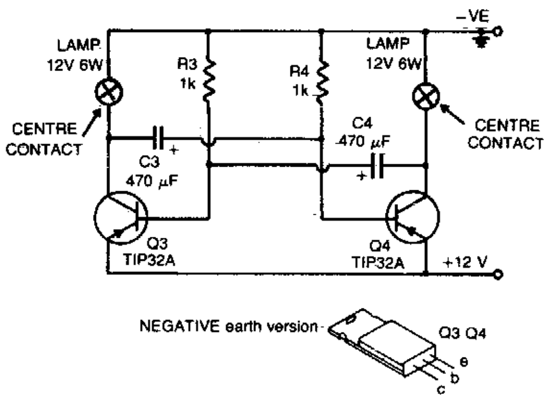


Fig. 34-1

Circuit Notes

Because of its ability to withstand the heavy inrush currents, this incandescent lamp flasher uses the C106 SCR. With the components shown, the flash rate is adjustable by potentiometer R3 within the range of 36 flashes per minute to 160 flashes per minute.

FLIP-FLOP FLASHER



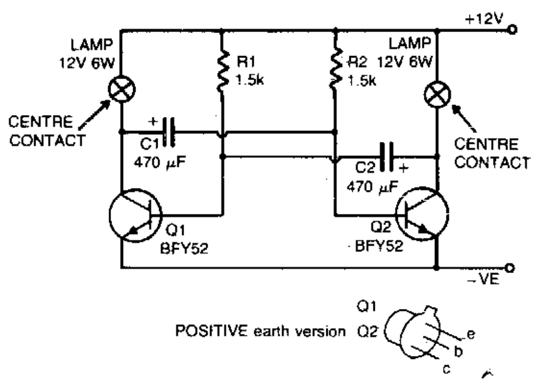
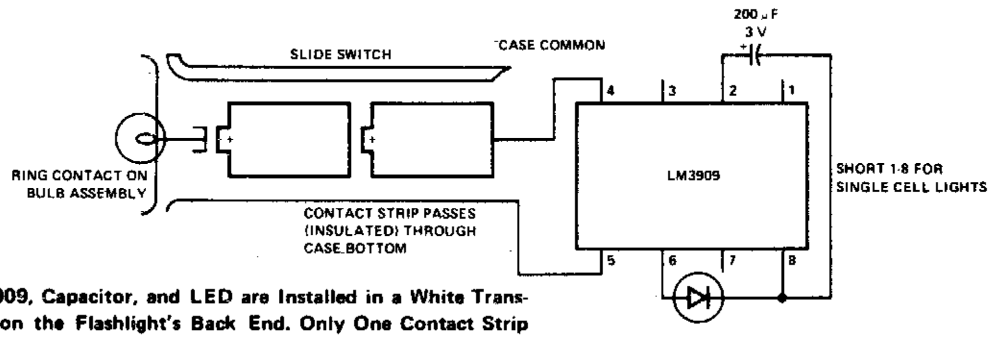


Fig. 34-2

Circuit Notes

The flashing action is provided by a simple astable multivibrator timed to give a flashing rate of about 60 flashes for each lamp per minute. Circuit for positive earth systems uses NPN transistors. The other uses PNP transistors.

FLASHLIGHT FINDER



Note: LM3909, Capacitor, and LED are Installed in a White Translucent Cap on the Flashlight's Back End. Only One Contact Strip (in Addition to the Case Connection) is Needed for Flasher Power. Drawing Current Through the Bulb Simplifies Wiring and Causes Negligible Loss Since Bulb Resistance Cold is Typically Less than 2 Ω.

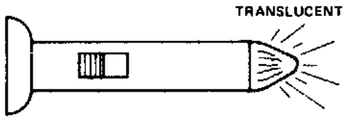
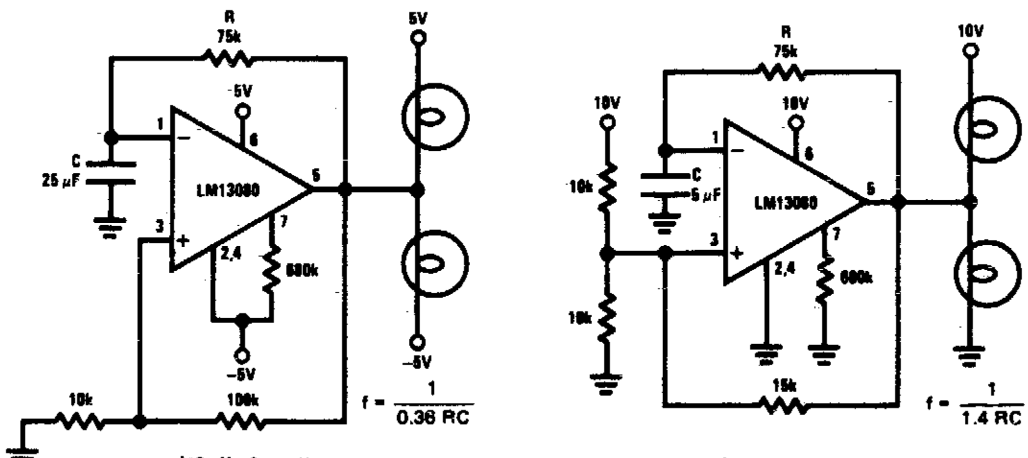


Fig. 34-3

Note: Winking LED Inside, Locates Light in Total Darkness

LOW FREQUENCY LAMP FLASHER/RELAY DRIVER



a) Split Supplies **b) Single Supply**

Fig. 34-4

Circuit Notes

This circuit is a low frequency warning device. The output of the oscillator is a square wave that is used to drive lamps or small relays. The circuit alternately flashes two incandescent lamps.

LOW COST RING COUNTER

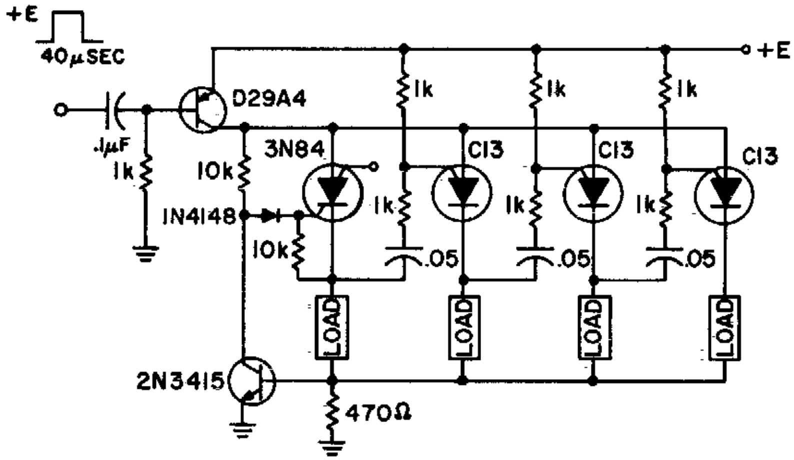


Fig. 34-5

Circuit Notes

This ring counter makes an efficient, low cost circuit featuring automatic resetting via the first stage 3N84. As many stages as desired may be cascaded.

RING COUNTER FOR INCANDESCENT LAMPS

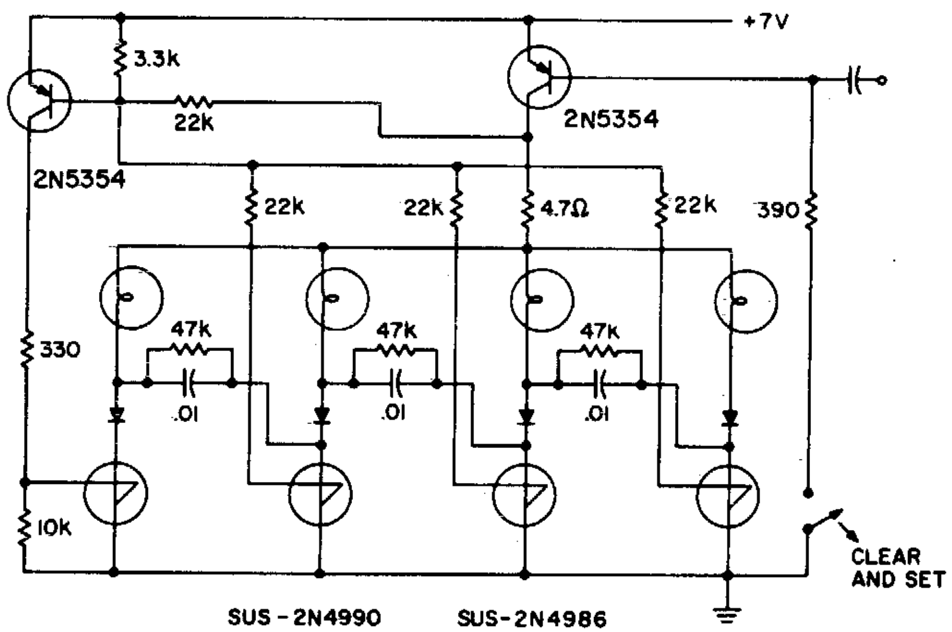


Fig. 34-6

DUAL LED CMOS FLASHER

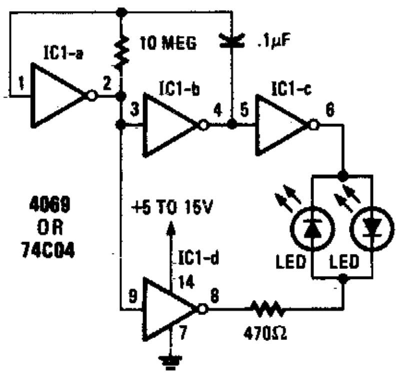


Fig. 34-7

Circuit Notes

Inverters IC1-a and IC1-b form a multivibrator and IC1-c is a buffer. Inverter IC1-d is connected so that its output is opposite that of IC1-c; when pin 6 is high, then pin 8 is low and vice versa. Because pins 6 and 8 are constantly changing state, first one LED and then the other is on since they are connected in reverse. The light seems to jump back and forth between the LED's. The 470-ohm resistor limits LED current. Depending upon the supply voltage used, the value of the resistor may have to be changed to obtain maximum light output. To change the switching rate, change the value of the capacitor.

AUTOMATIC SAFETY FLASHER

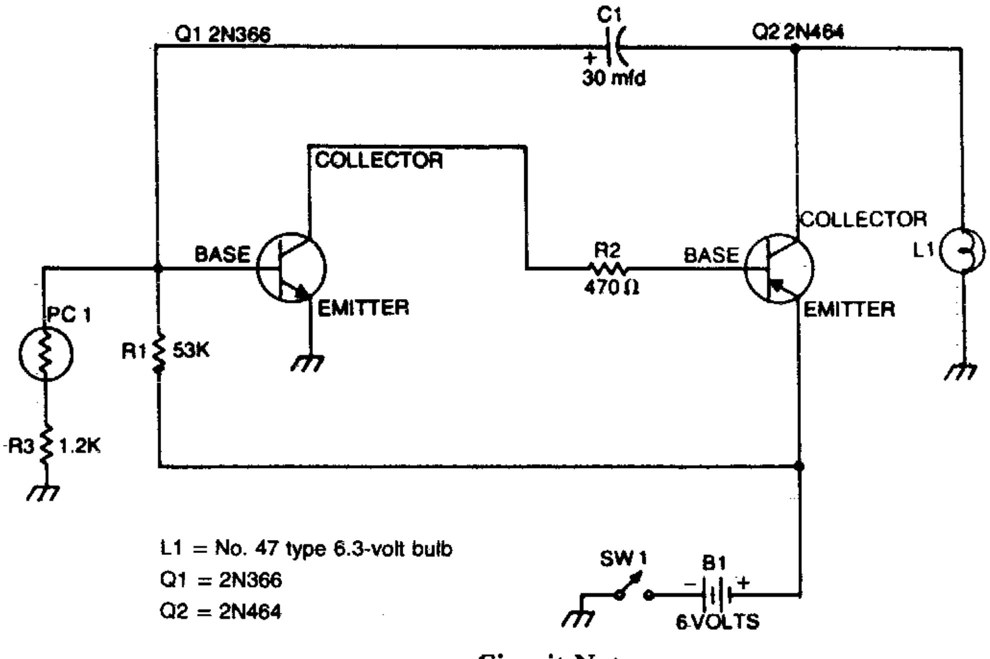


Fig. 34-8

Circuit Notes

This flasher only comes on at night. It furnishes a bright nighttime illumination, and shuts itself off automatically as soon as the sun comes up. The photocell must be mounted on top of the unit in such a way as to detect the greatest amount of available light.

NEON BLINKER

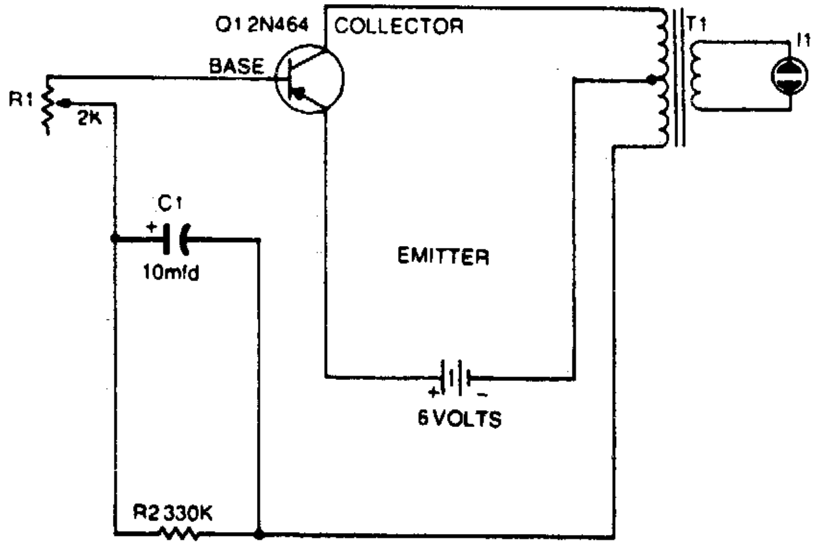
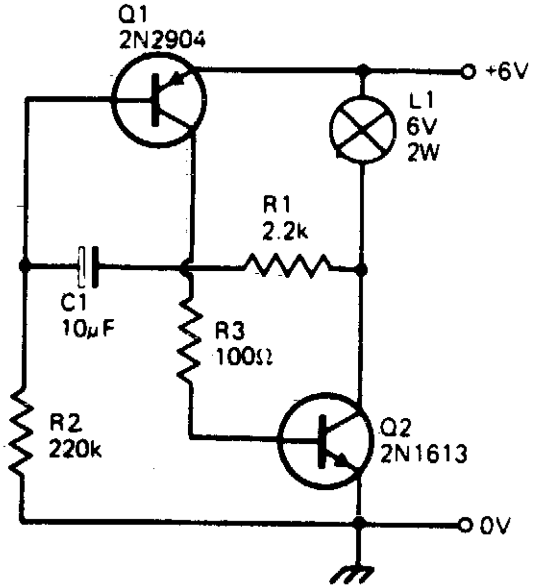


Fig. 34-9

Circuit Notes

The universal output transformer and the transistor form a low-frequency oscillator. The rate of flashing of the neon bulb is determined by potentiometer R1.

TRANSISTORIZED FLASHER



Circuit Notes

This simple circuit will flash a 6 volt lamp at a rate determined by the size of capacitor C1. It is most economical on power as it only draws current when the lamp is on. When the lamp is off, both transistors are biased off.

Fig. 34-10

FLASHER/LIGHT CONTROL

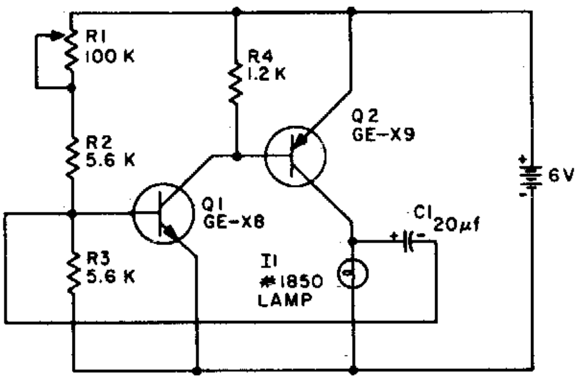


Fig. 34-11

Parts List

C1 — 20-mfd, 6-volt electrolytic capacitor
I1 — 6-volt, GE No. 1850 lamp and socket
Q1 — GE-X8 transistor
Q2 — GE-X9 transistor
R1 — 100K-ohm, 2-watt potentiometer
R2, R3 — 5.6K-ohm, 1/2-watt resistor
R4 — 1.2K-ohm, 1/2-watt resistor
Battery — 6-volt dry pack

Circuit Notes

The circuit is a two-stage, direct-coupled transistor amplifier connected as a free-running multivibrator. Both the flash duration and flash interval can be changed by turning the potentiometer, R1.

NEON TUBE FLASHER

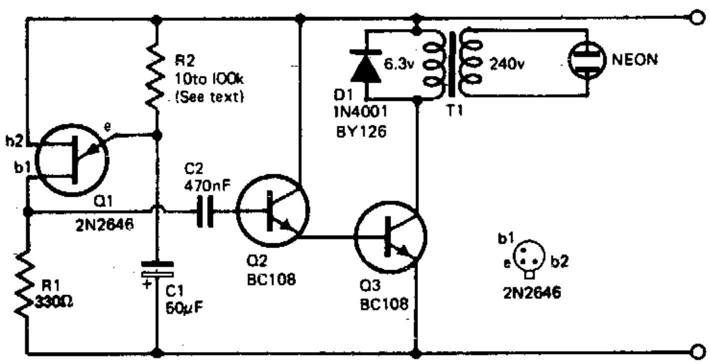


Fig. 34-12

Circuit Notes

The voltage required to ignite the neon tube is obtained by using an ordinary filament transformer (240-6.3 V) in reverse. Battery drain is quite low, around 1 to 2 milliamps for a nine volt battery. The pulses from Q1, unijunction transistor, operated as a relaxation oscillator and are applied to Q2 which in turn drives Q3 into saturation. The sharp rise in current through the 6.3 V winding of the transformer as Q3 goes into saturation induces a high voltage in the secondary winding causing the neon to flash. The diode D1 protects the transistor from high voltage spikes generated when switching currents in the transformer.

DC FLASHER WITH ADJUSTABLE ON AND OFF TIME

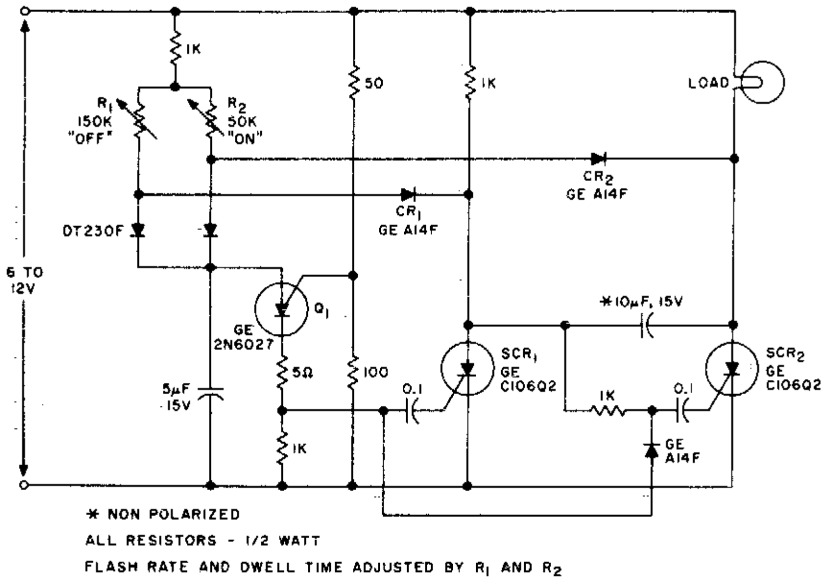


Fig. 34-13

Circuit Notes

This circuit utilizes a power flip-flop and programmable unijunction (PUT) to obtain adjustable on and off times.

LOW VOLTAGE FLASHER

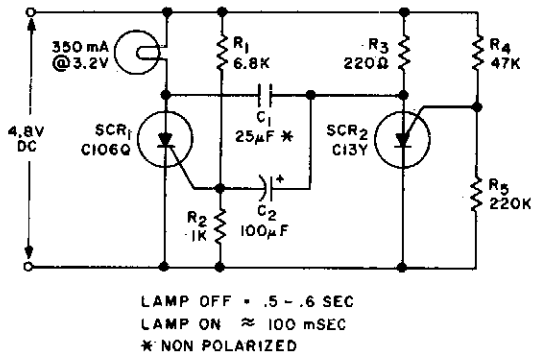


Fig. 34-14

Circuit Notes

Applying voltage to the circuit triggers SCR1. With SCR1 on, the voltage on the anode of SCR2 rises until SCR2 triggers to commutate SCR1. The voltage on the gate of SCR1 will swing negative at this time, and only after a positive potential of ≈ 0.5 volt is once again attained, will SCR1 retrigger. The circuit could be used for higher voltage levels, but the peak negative voltage on the gate of SCR1 must be limited to less than 6 volts.

1 A LAMP FLASHER

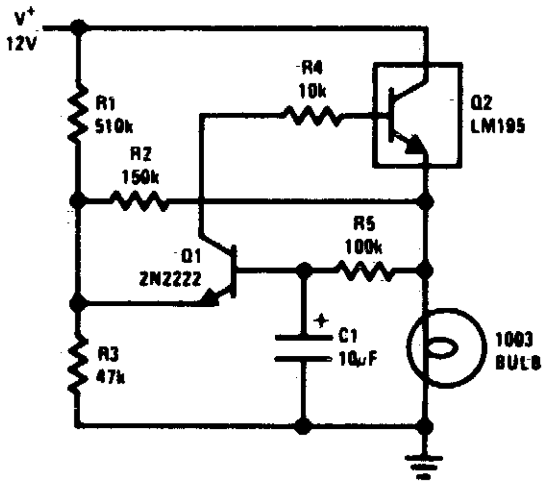
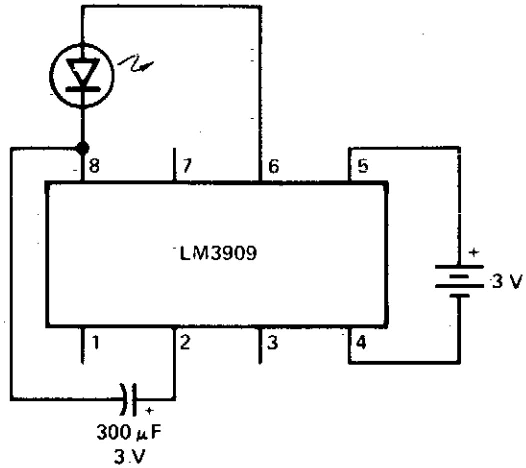


Fig. 34-15

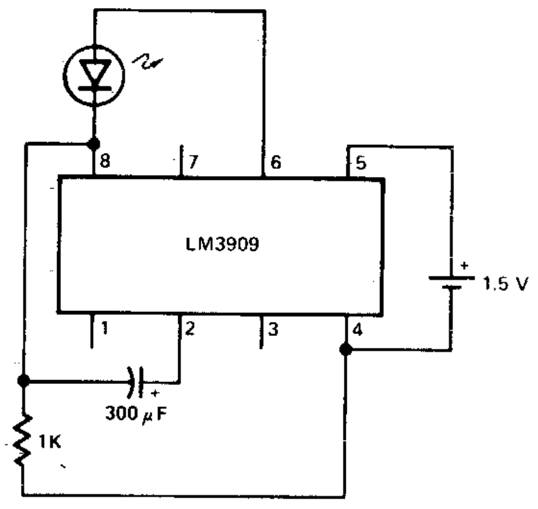
3 V FLASHER



Note: Nominal Flash Rate: 1 Hz. Average I_{DRAIN} = 0.77 mA

Fig. 34-17

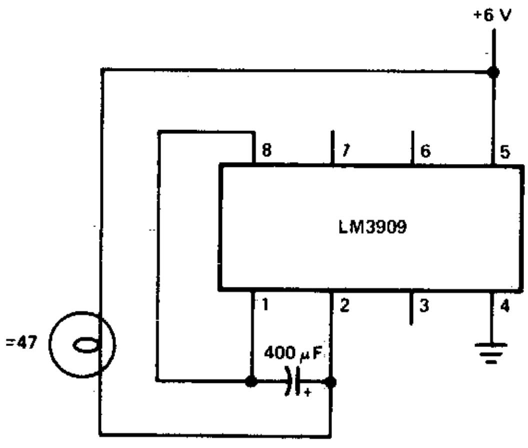
FAST BLINKER



Note: Nominal Flash Rate: 2.6 Hz. Average I_{DRAIN} = 1.2

Fig. 34-16

INCANDESCENT BULB FLASHER



Note: Flash Rate: 1.5 Hz

Fig. 34-18

FLASHER FOR 4 PARALLEL LEDs

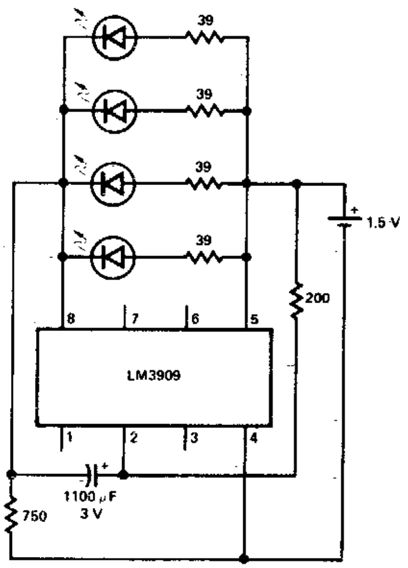


Fig. 34-19

Note: Nominal Flash Rate:
1.3 Hz. Average I_{DRAIN} = 2 mA

SAFE, HIGH VOLTAGE FLASHER

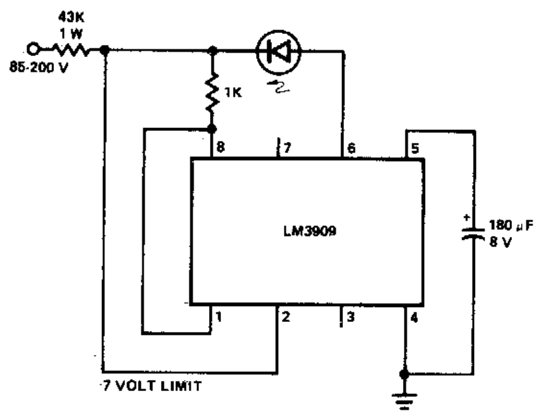
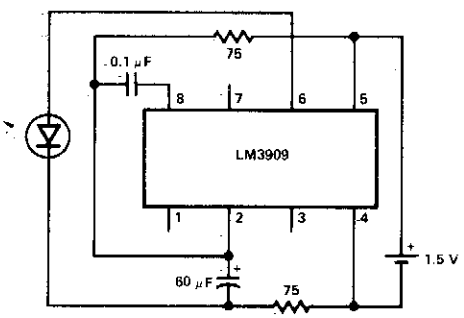


Fig. 34-21

LED BOOSTER



Note: High efficiency, 4 mA drain

Note: Continuous Appearing Light Obtained By Supplying Short, High Current, Pulses (2 kHz) to LEDs With Higher Than Battery Voltage Available.

Fig. 34-20

ALTERNATING FLASHER

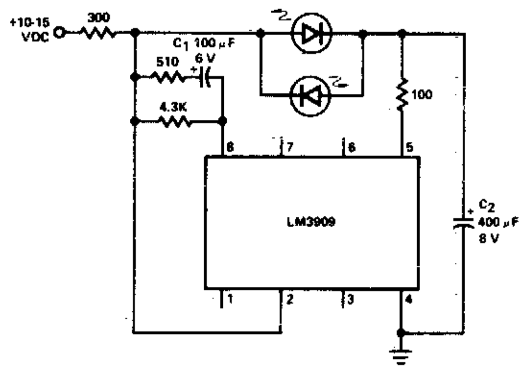
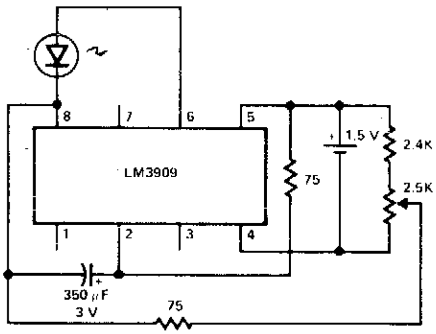


Fig. 34-22

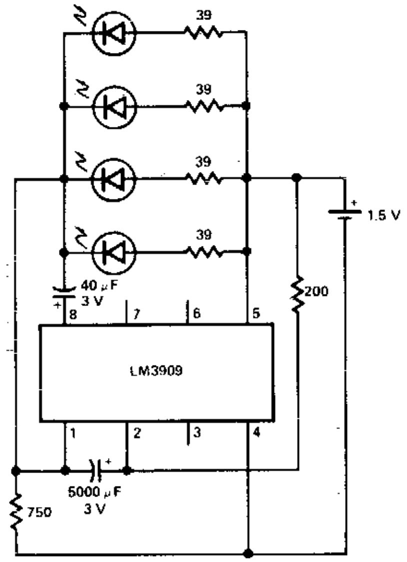
VARIABLE FLASHER



Note: Flash Rate: 0-20 Hz

Fig. 34-23

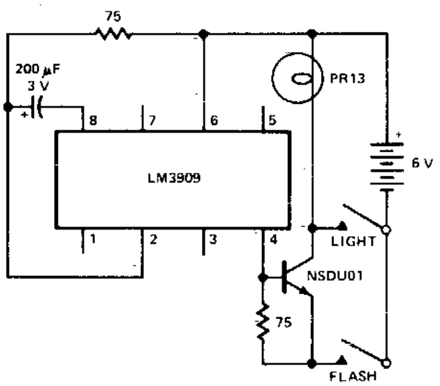
HIGH EFFICIENCY PARALLEL CIRCUIT FLASHER



Note: Nominal Flash Rate: 1.5 Hz. Average I_{DRAIN} = 1.5 mA

Fig. 34-25

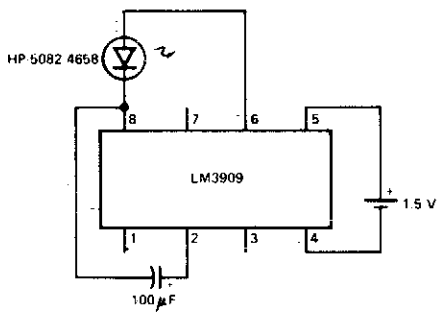
EMERGENCY LANTERN/FLASHER



Note: Nominal Flash Rate: 1.5 Hz

Fig. 34-24

MINIMUM POWER FLASHER (1.5 V)



Note: Nominal Flash Rate: 1.1 Hz. Average I_{DRAIN} = 0.32 mA

Fig. 34-26

35

Frequency Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

INEXPENSIVE FREQUENCY COUNTER/TACHOMETER

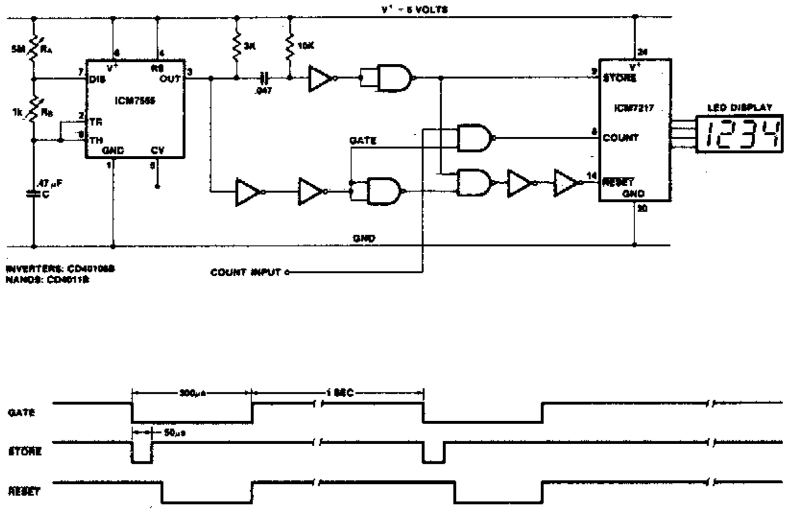


Fig. 35-1

Circuit Notes

This circuit uses the low power ICM7555 (CMOS 555) to generate the gating, STORE and RESET signals. To provide the gating signal, the timer is configured as an astable multivibrator. The system is calibrated by using a 5 M potentiometer for R_A as a coarse control and a 1 jk potentiometer for R_B as a fine control. CD40106B's are used as a monostable multivibrator and reset time delay.

LINEAR FREQUENCY METER (AUDIO SPECTRUM)

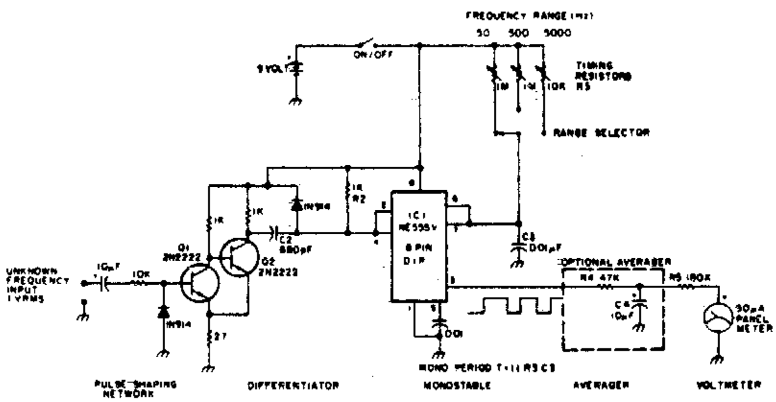


Fig. 35-2

Circuit Notes

The 555 is used in a monostable multivibrator circuit that puts out a fixed timewidth pulse, which is triggered by the unknown input frequency.

POWER-LINE FREQUENCY METER

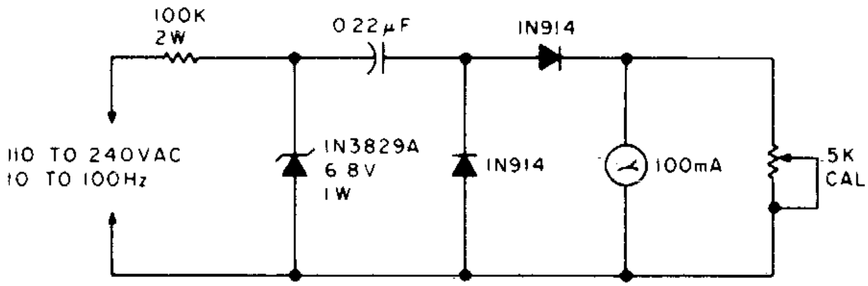


Fig. 35-3

Circuit Notes

The meter will indicate the frequency from a power generator. Incoming sine waves are converted to square waves by the 100 K resistor and the 6.8 V zener. The square wave is differentiated by the capacitor and the current is averaged by the diodes. The average current is almost exactly proportional to the frequency and can be read directly on a 100 mA meter. To calibrate, hook the circuit up to a 60 Hz poweraline and adjust the 5 K pot to read 60 mA.

AUDIO FREQUENCY METER

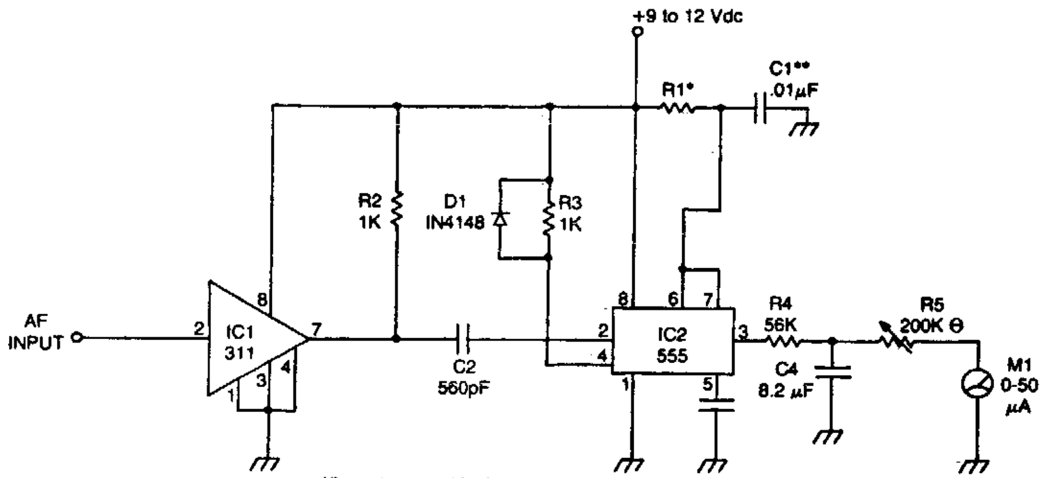


Fig. 35-4

*R1: 1 meg 100K 10K
RANGE: 50 500 5000
(Hz)
**C1 : POLYSTYRENE OR SILVER-MICA

Circuit Notes

The meter uses time averaging to produce a direct current that is proportional to the frequency of the input signal.

36

Frequency Multipliers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

BROADBAND FREQUENCY DOUBLER

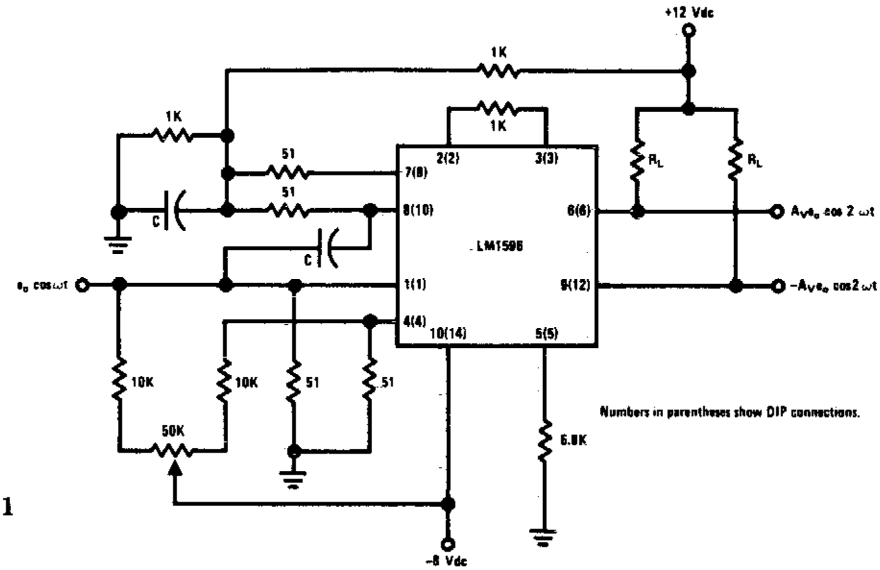


Fig. 36-1

Circuit Notes

This circuit will double low-level signals with low distortion. The value of C should be chosen for low reactance at the operating frequency. Signal level at the carrier input must be less than 25 mV peak to maintain operation in the linear region of the switching differential amplifier. Levels to 50 mV peak may be used with some distortion of the output waveform. If a larger input signal is available, a resistive divider may be used at the carrier input with full signal applied to the signal input.

FREQUENCY DOUBLER

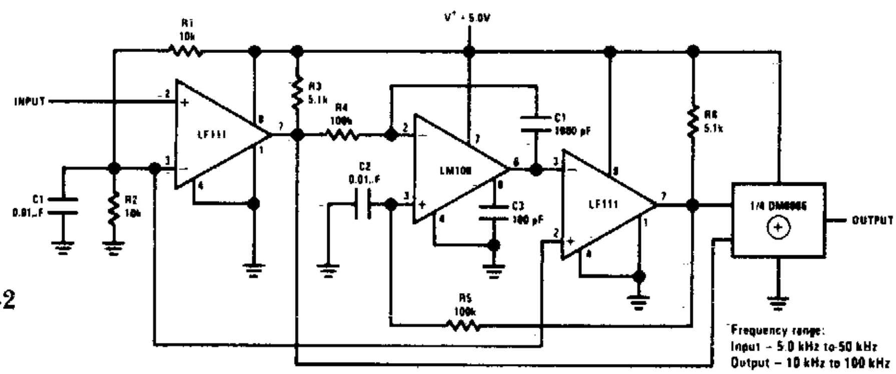


Fig. 36-2

150 TO 300 MHz DOUBLER

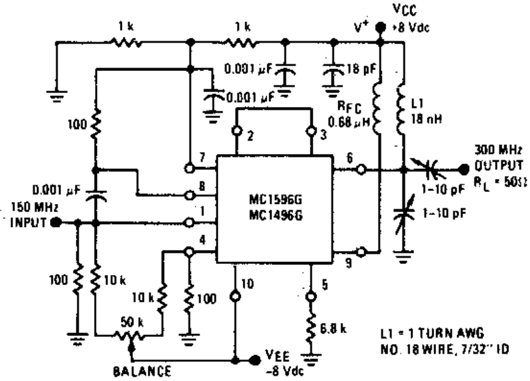


Fig. 36-3

LOW-FREQUENCY DOUBLER

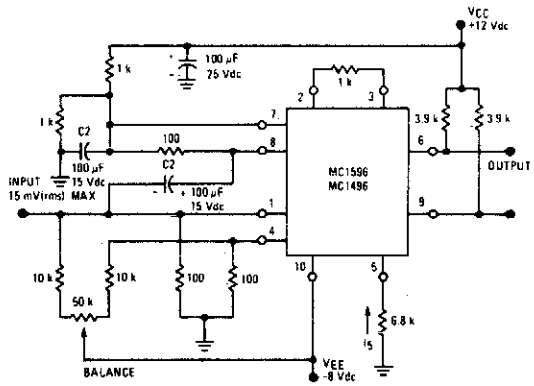


Fig. 36-4

OSCILLATOR WITH DOUBLE FREQUENCY OUTPUT

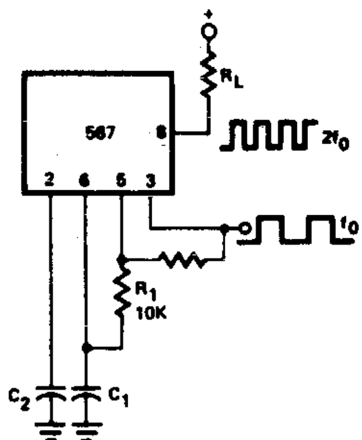


Fig. 36-5

37

Frequency-to-Voltage Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DC-10 kHz FREQUENCY/VOLTAGE CONVERTER

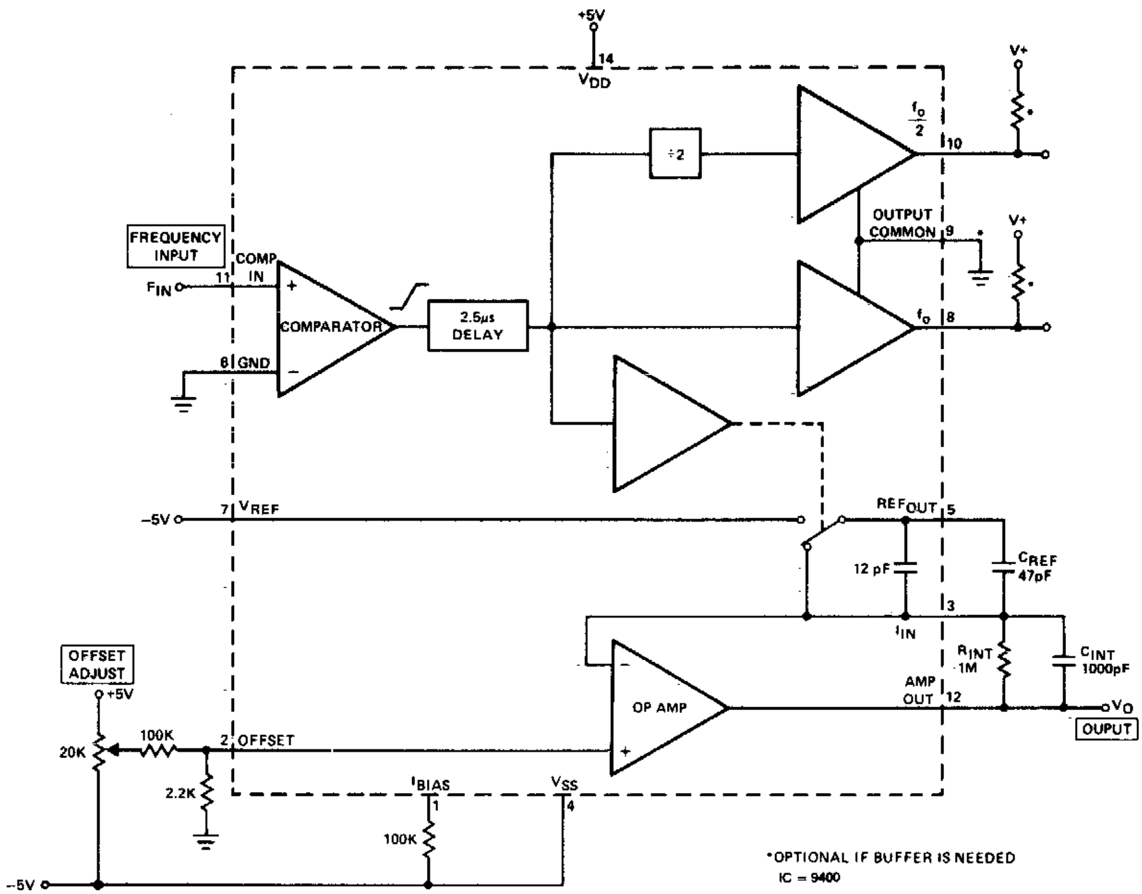
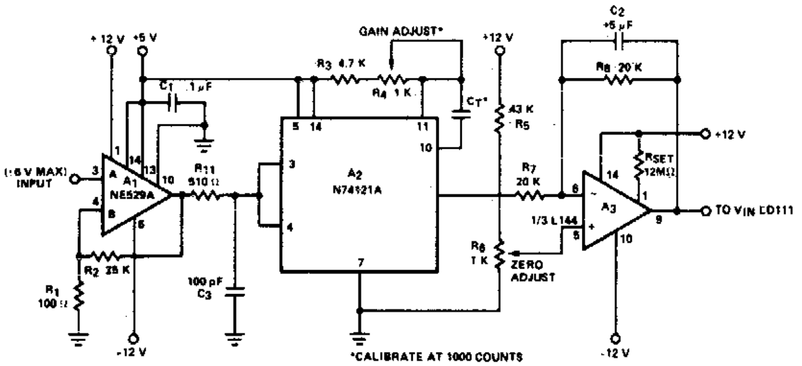


Fig. 37-1

Circuit Notes

The converter generates an output voltage which is linearly proportional to the input frequency waveform. Each zero crossing at the comparator's input causes a precise amount of change to be dispensed into the op amp's summing junction. This charge in turn flows through the feedback resistor generating voltage pulses at the output of the op amp. Capacitor (C_{INT}) across R_{INT} averages these pulses into a dc voltage which is linearly proportional to the input frequency.

FREQUENCY-TO-VOLTAGE CONVERTER (DIGITAL FREQUENCY METER)



RANGE	C_T
2 kHz	.082 μF
20 kHz	.0082 μF
200 kHz	820 pF
2 MHz	82 pF
20 MHz	8.2 pF

Fig. 37-2

Circuit Notes

This circuit converts frequency to voltage by taking the average dc value of the pulses from the 74121 monostable multivibrator. The one shot is triggered by the positive-going ac signal at the input of the 529 comparator. The amplifier acts as a dc filter, and also provides zeroing. The accuracy is 2% over a 5 decade range. The input signal to the comparator should be greater than 0.1 volt peak-to-peak, and less than 12 volts peak-to-peak for proper operation.

ZENER REGULATED FREQUENCY-TO-VOLTAGE CONVERTER

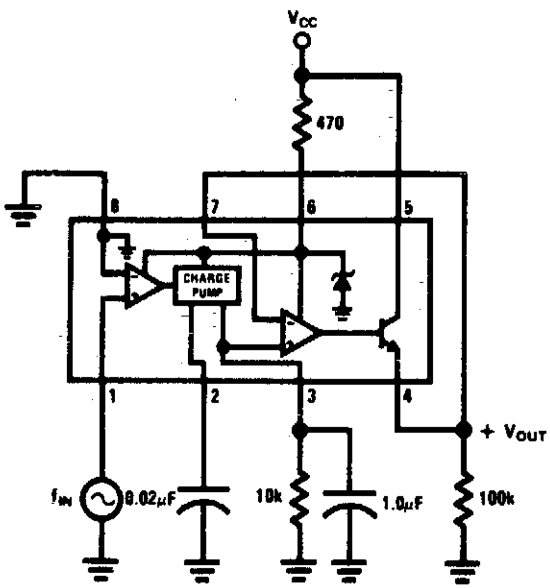
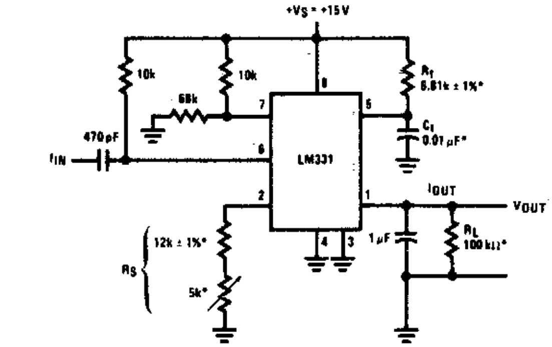


Fig. 37-3

SIMPLE FREQUENCY-TO-VOLTAGE CONVERTER (10 kHz FULL-SCALE, ±0.006% NON-LINEARITY)

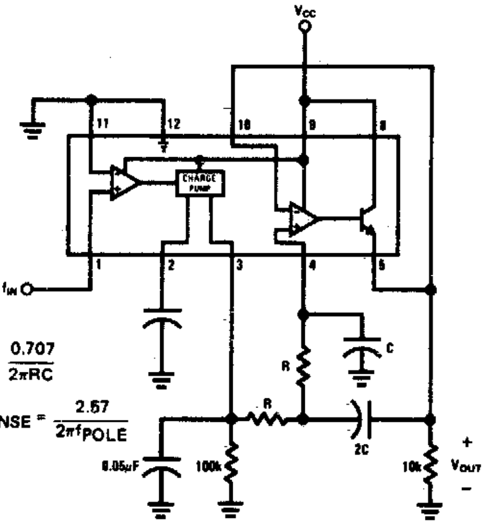


$V_{OUT} = f_{IN} \times 2.09V \times \frac{R_L}{R_S} \times (R_tC_t)$

*Use stable components with low temperature coefficients.

Fig. 37-4

FREQUENCY-TO-VOLTAGE CONVERTER WITH 2-POLE BUTTERWORTH FILTER TO REDUCE RIPPLE



$f_{POLE} = \frac{0.707}{2\pi RC}$

$\tau_{RESPONSE} = \frac{2.57}{2\pi f_{POLE}}$

Fig. 37-6

F/V CONVERSION, TTL INPUT

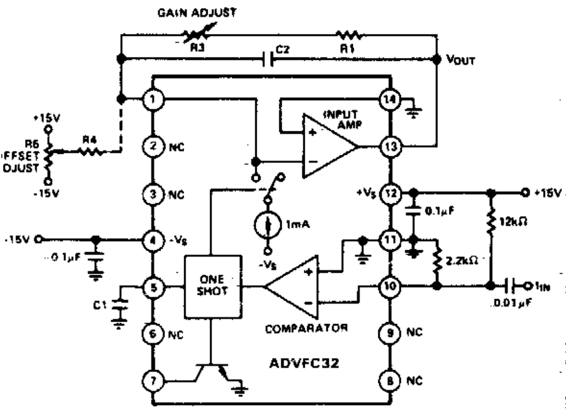
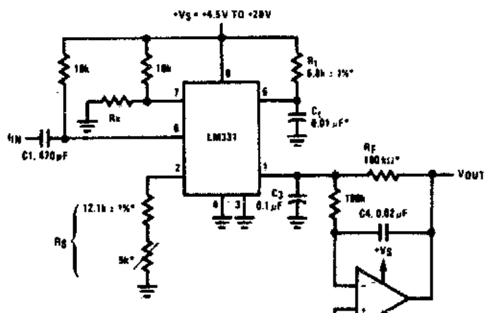


Fig. 37-5

PRECISION FREQUENCY-TO-VOLTAGE CONVERTER (10 kHz FULL/SCALE WITH 2-POLE FILTER, ±0.01% NON-LINEARITY MAXIMUM)



$V_{OUT} = -f_{IN} \times 2.09V \times \frac{R_F}{R_S} \times (R_tC_t)$

$\text{SELECT } R_X = \frac{(V_S - 2V)}{0.2mA}$

*Use stable components with low temperature coefficients.

Fig. 37-7

38

Fuzz Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FUZZ BOX 1

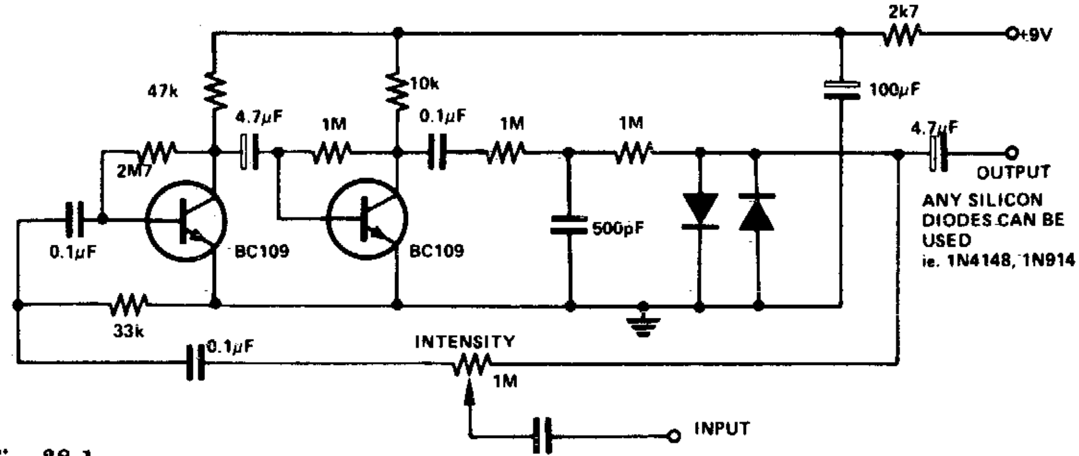
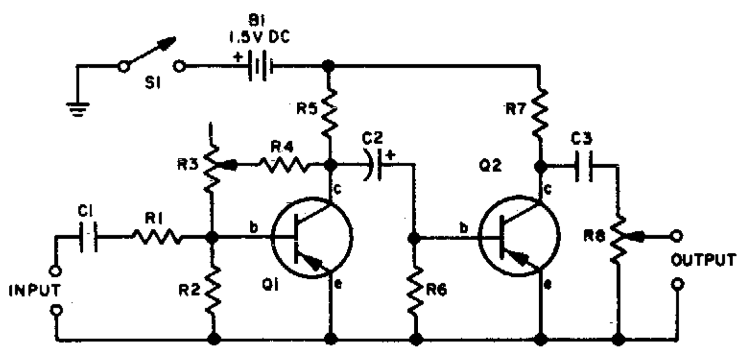


Fig. 38-1

Circuit Notes

The input signal is amplified by the transistors. The distorted output is then clipped by the two diodes and the high frequency noise is filtered from the circuit via the 500 pF capacitor. The 1 M pot adjusts the intensity of the fuzz from maximum to no fuzz (normal playing).

FUZZ BOX 2



- B1—1.5-V AA battery
- C1, C3—0.1-uF, 50-VDC capacitor
- C2—4.7-uF, 10-VDC electrolytic capacitor
- Q1, Q2—pnp transistor—HEP-632
- R1, R6—22,000-ohm, ½-watt resistor
- R2—18,000-ohm, ½-watt resistor
- R3—1-megohm pot
- R4—100,000-ohm, ½-watt resistor
- R5, R7—10,000-ohm, ½-watt resistor
- R8—50,000-ohm pot
- S1—Spst switch

Fig. 38-2

Circuit Notes

Potentiometer R3 sets the degree of fuzz, and R8 sets the output level. Since the fuzz effect cannot be completely eliminated by R3, fuzz-free sound requires a bypass switch from the input to output terminals.

FUZZ BOX 3

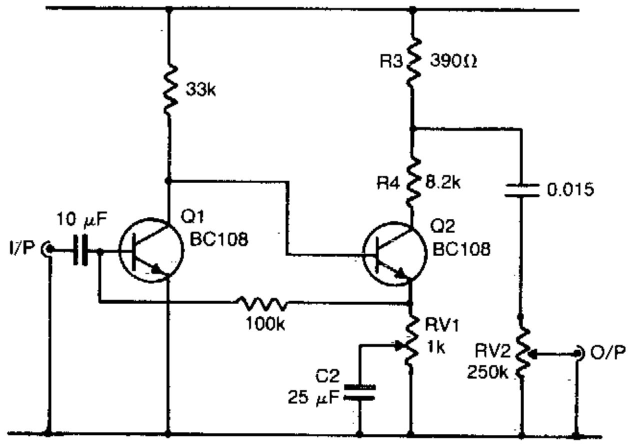


Fig. 38-3

Circuit Notes

Q1 and Q2 form a voltage amplifier which has sufficient gain to be overdriven by a relatively low input, such as an electric guitar. The result is that the output from Q2 is a Squared-Off verson of the input, giving the required fuzz sound. RV1 adjusts the amount of negative feedback inserted into the circuit by C2, and thus the amount of squaring of the signal. The purpose of R3 and R4 is to lower the output voltage to a suitable level, which is then adjusted as required with the volume control VR2.

FUZZ BOX 4

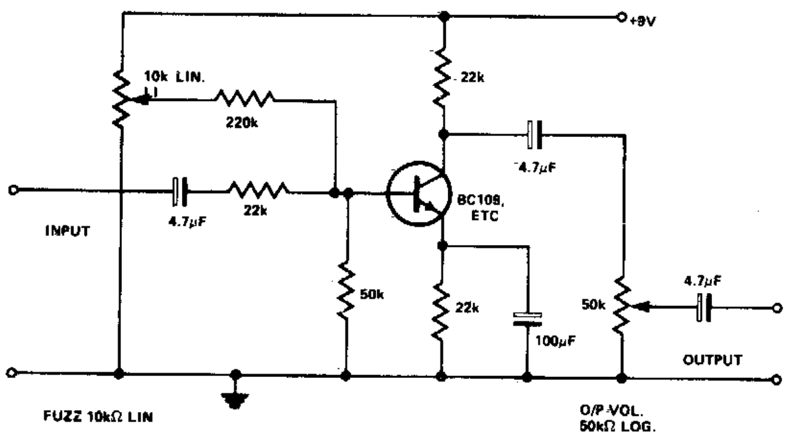
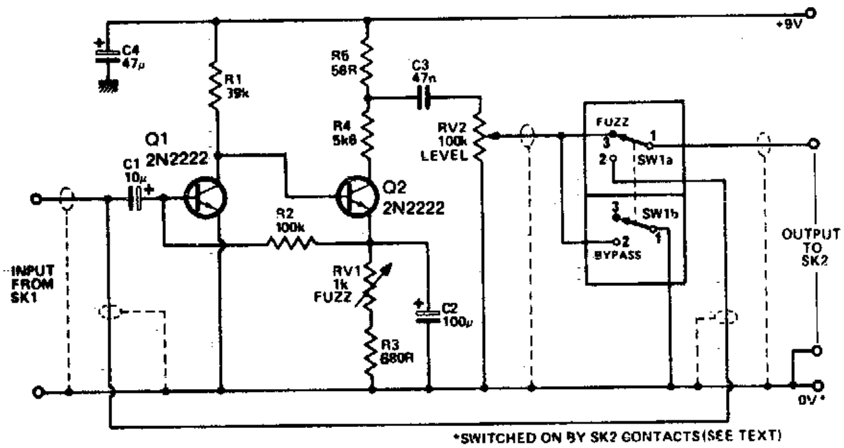


Fig. 38-4

Circuit Notes

None of the components are particularly critical in value or quality, as distortion is the sole object! The transistor could be BC107-8-9, 2N2926, etc.

FUZZ BOX 5



Circuit Diagram

Fig. 38-5

Circuit Notes

Transistors Q1 and Q2 amplify the incoming signal, and the gain is such that the input will overload when used with an electric guitar. RV1 adjusts the amount of feedback present, and hence voltage gain. The output is, therefore, a squared version of the input signal. The amount of squaring is varied by RV1.

GUITAR FUZZ

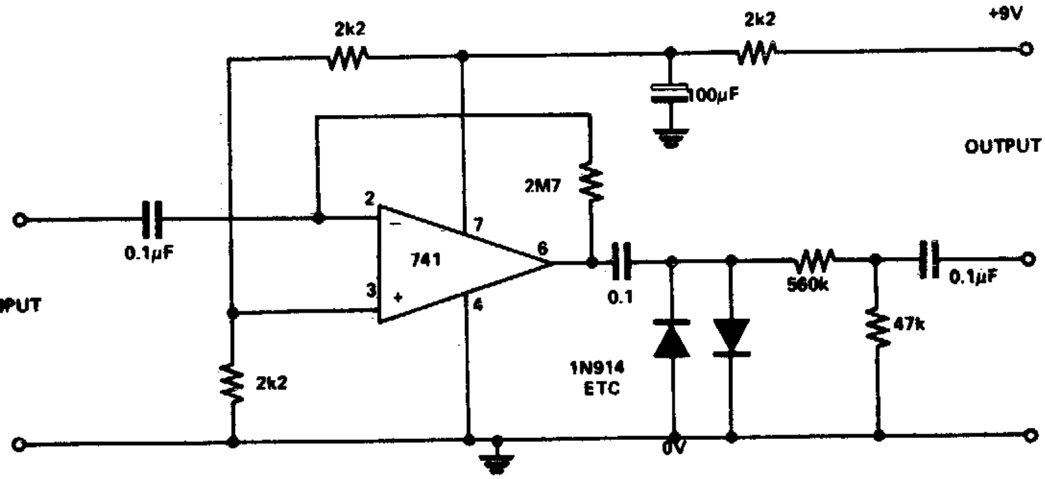


Fig. 38-6

Circuit Notes

The 741 has a maximum gain of 20,000, but the circuit is so designed that the IC's gain is 2,700,000 which then distorts the output. This distortion gives the fuzz effect. The two diodes clip the output to drop the level, also lowered by the potential divider. This circuit also sustains the notes, due to clipping, giving a totally new sound.

39

Games

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Ready, Set, Go!
- Electronic Dice
- Game Roller or Chase Circuit
- Toss-A-Coin Binary Box
- Electronic Coin Tosser
- Heads or Tails
- Pot Shot
- Low Cost Heads or Tails
- Who Is First
- Windicator

READY, SET, GO!

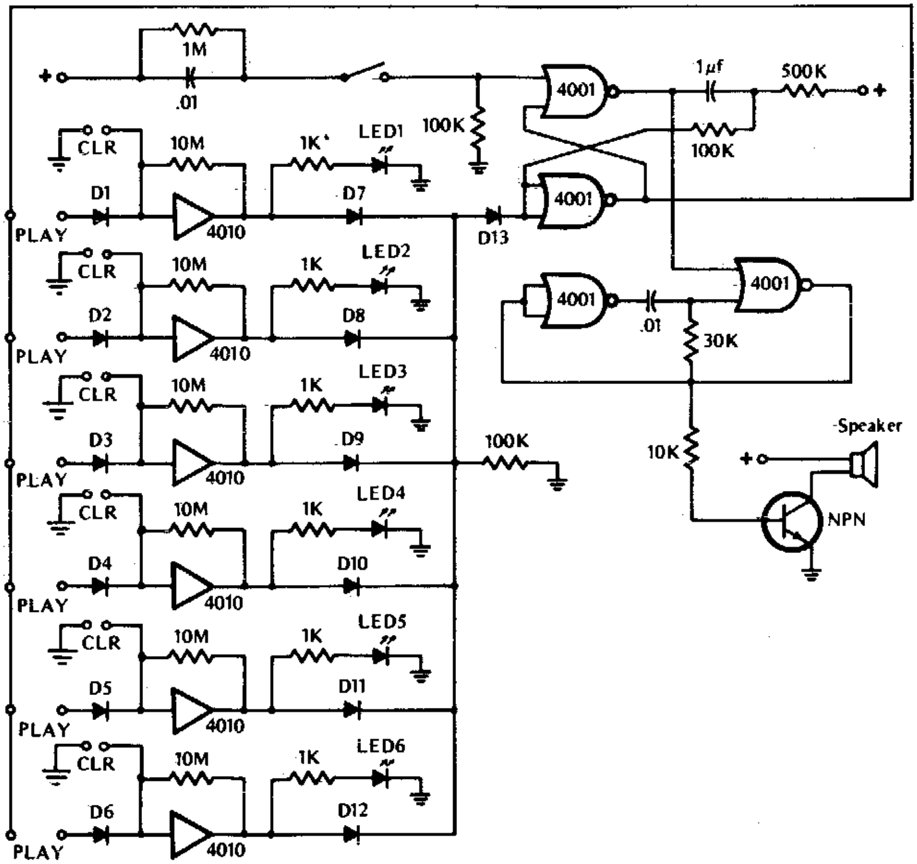


Fig. 39-1

Circuit Notes

This game tests a player's reaction time. It is activated by closing switch S1, which starts the tone generator and arms the circuit. The touchplate, labeled PLAY in the diagram, consists of two metal strips about 1/16th-inch apart. The first player to bridge the gap with his or her finger turns off the tone and lights the associated LED indicator. A second touchplate, labeled CLR in the diagram, clears the circuit, extinguishing the LED, when its gap is bridged by a fingertip.

ELECTRONIC DICE

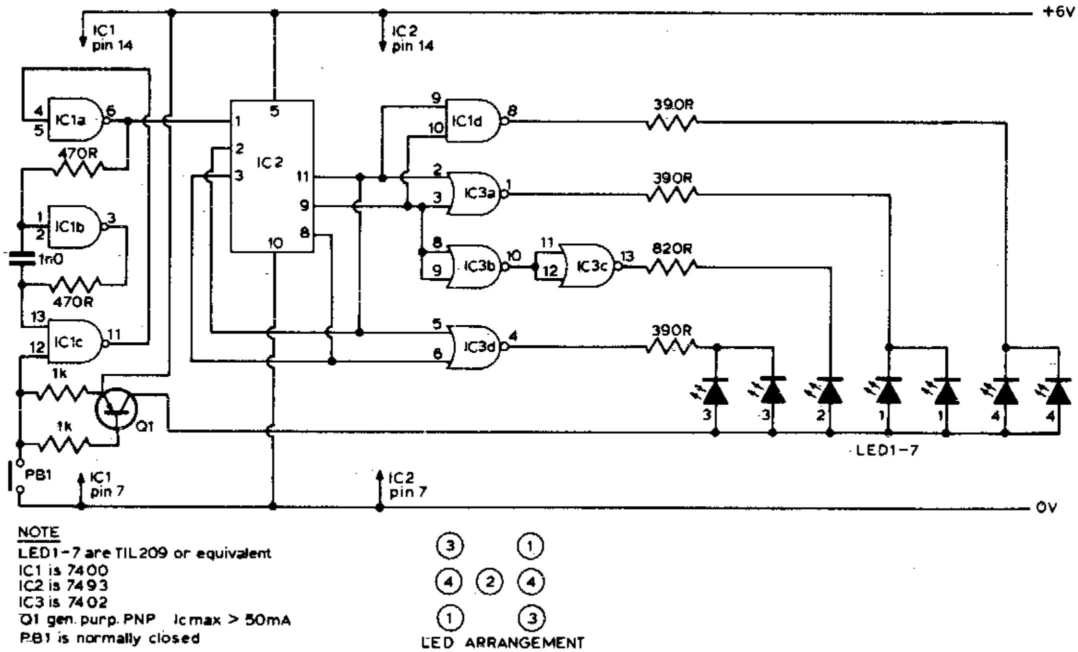


Fig. 39-2

Circuit Notes

Six LEDs are arranged to produce a display the same as the dots on a dice. When PBI is depressed, the display is blanked and the oscillator (IC1 a, b, c) clocks IC2 at about 1MHz. IC2 counts from zero and resets on seven. When PBI is released, the display is enabled and a decoding system (IC3) produces the correct output on the LEDs.

GAME ROLLER OR CHASE CIRCUIT

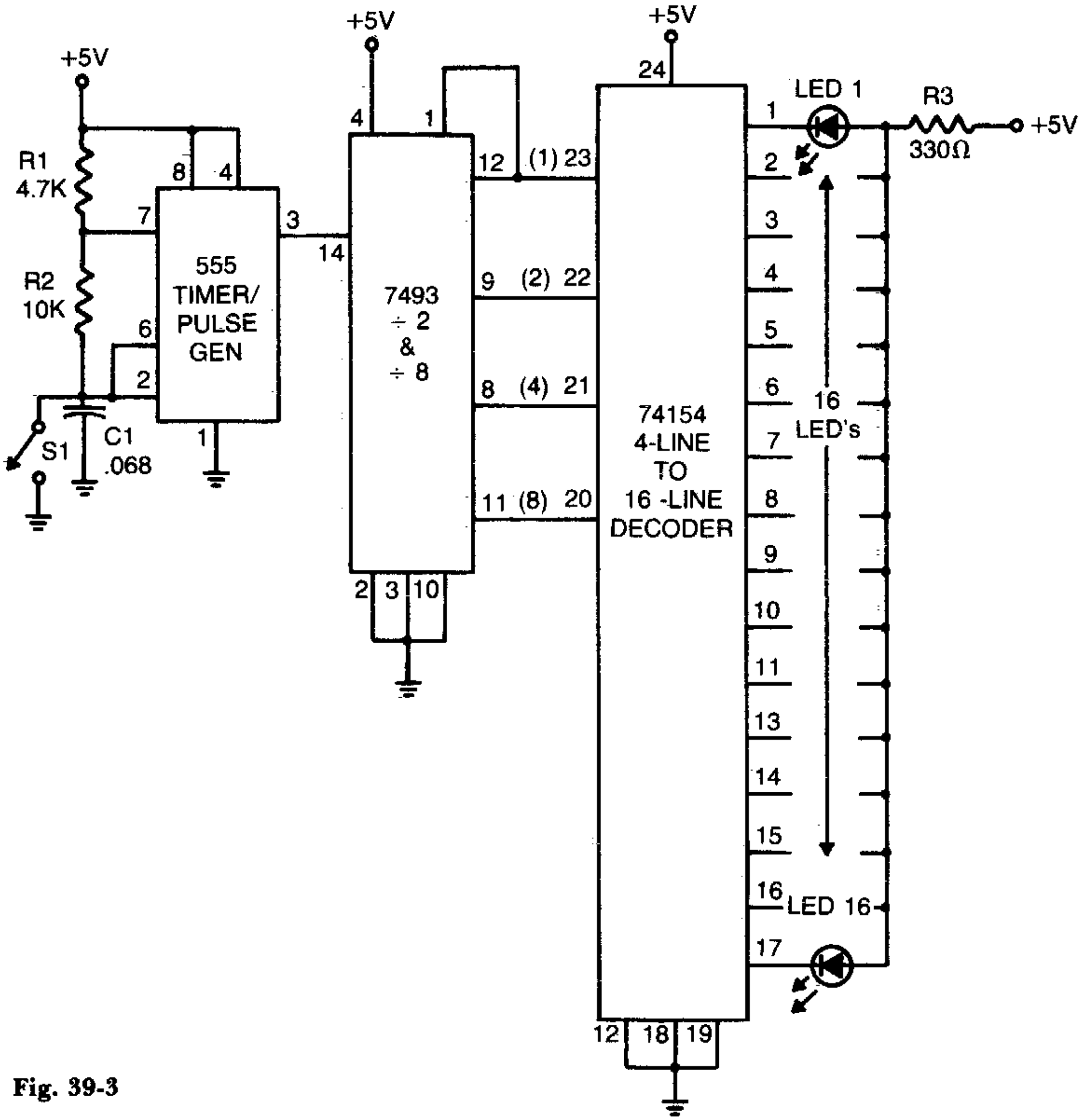


Fig. 39-3

Circuit Notes

The 555 timer produces a rapid series of pulses whenever switch S1 is open. These pulses are counted in groups of 16 and converted into binary form by the 7493 and applied to the 74154 (a 1-of-16 decoder/demultiplexer) wired so that each of its 16 output lines goes low sequentially and in step with the binary count delivered by the 7493. When the switch is closed, only one LED remains on. Only one current limiting resistor (R3) is used for all the LED's since only one is on at any one time.

TOSS-A-COIN BINARY BOX

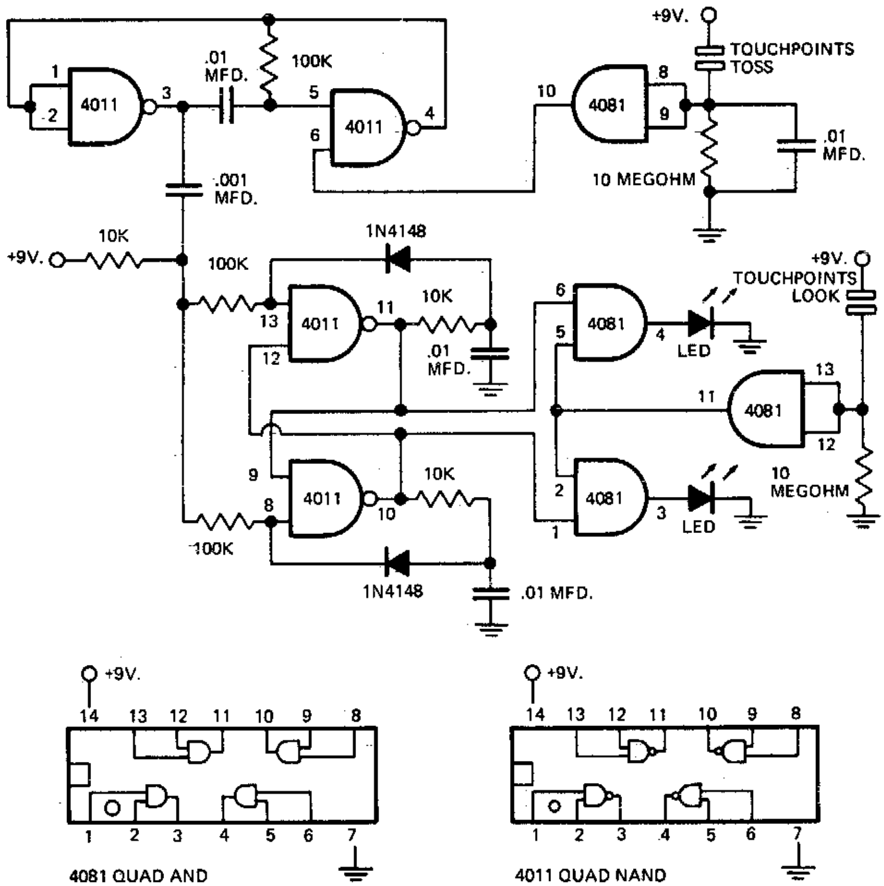


Fig. 39-4

Circuit Notes

Circuit uses an astable multivibrator to vary the heads-or-tails condition, and a flip-flop to store the condition given by the multivibrator. Consequently, the circuit is wired so that the flip-flop's state is changed once for each full cycle the multivibrator goes through to assure an absolutely even 50-50 chance of a heads or tails loss.

ELECTRONIC COIN TOSSER

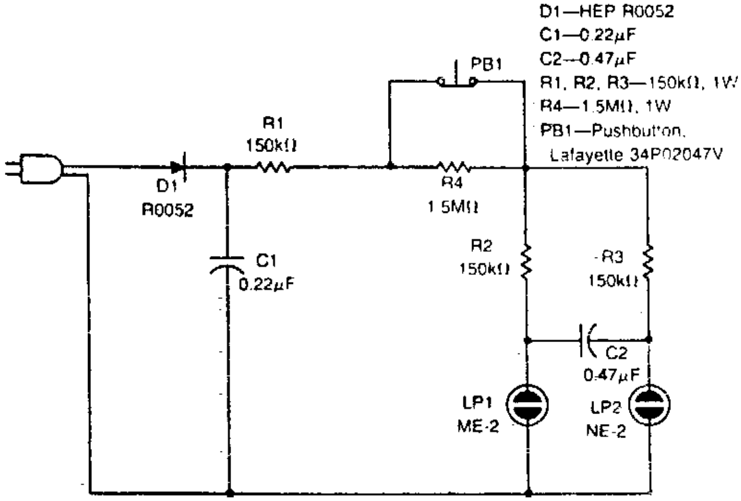


Fig. 39-5

Circuit Notes

The circuit shown simulates the flipping of a coin by merely pushing switch PB1.

HEADS OR TAILS

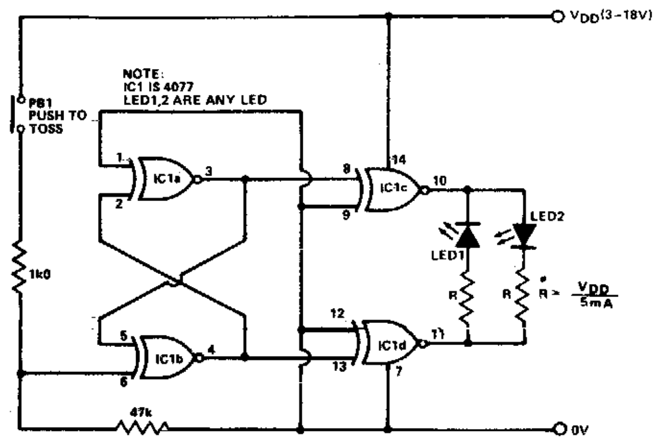


Fig. 39-6

Circuit Notes

This ultra-simple heads or tails indicator uses a single 4077 and no capacitor.

The circuit is normally in a latched bistable mode; when the switch is closed the circuit will oscillate, i.e. toss the coin. The astable frequency is approximately 5-10 MHz. PB1 is a normally closed switch.

POT SHOT

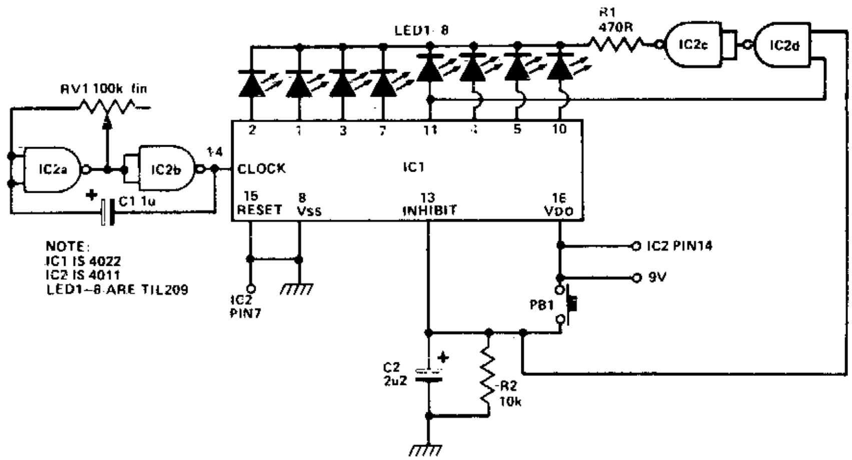


Fig. 39-7

Circuit Notes

This is a circuit for a game of the shooting gallery variety. IC2a and b form an astable multivibrator clocking IC1 which causes LEDs 1-8 to flash in turn LED 5 is the target LED and the object of the game is to depress PBI just as LED 5 comes on. If this is done, the whole display is blanked for a few seconds signifying a hit. Otherwise, the LED which was lit remains lit. When the push button is released, C2 discharges through R2 taking 8 pin 13 low again and the LEDs will start to flash again.

LOW COST "HEADS OR TAILS"

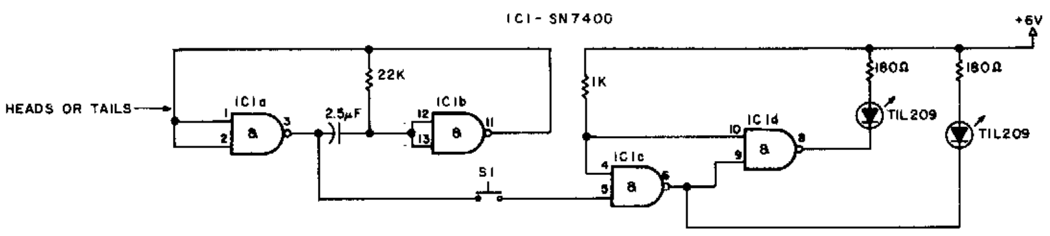


Fig. 39-8

Circuit Notes

S1 must be a push-to-make, release-to-break, switch.

WHO IS FIRST

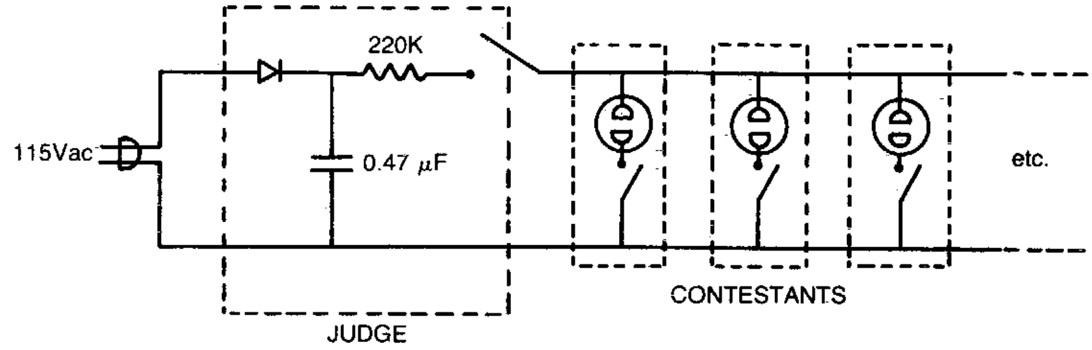


Fig. 39-9

Circuit Notes

Here is a circuit for any question-and-answer party game. The first button pushed unionizes the neon bulb dropping the dc voltage on the parallel neons (the other contestants) below the ionization level: determining unequivocally the first person to press the button.

WINDICATOR

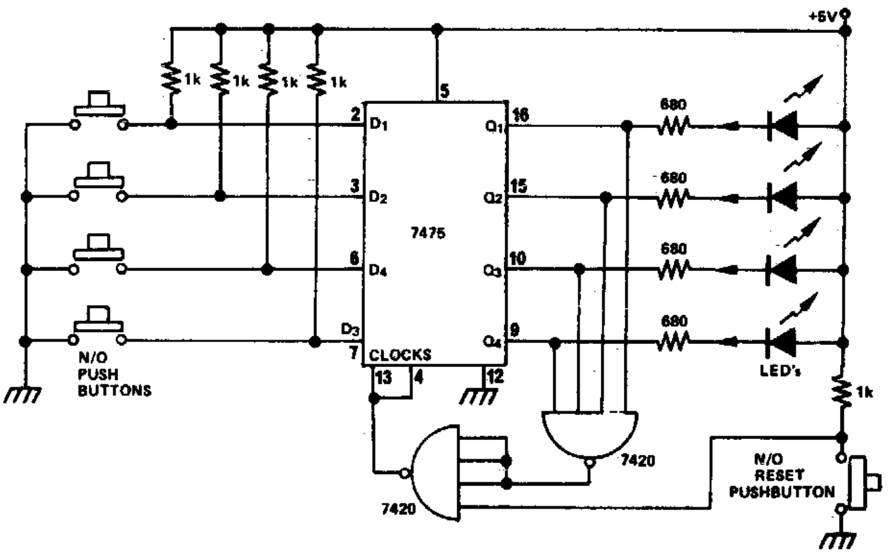


Fig. 39-10

Circuit Notes

Two TTL ICs and a handful of other components are all that is needed for a circuit that will indicate which of four buttons was pressed first, as well as lock out all other entries. A logic 0 at one of the Q outputs, lights the appropriate LED and locks out other entries by taking the clock input low.

40

Gas/Vapor Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

GAS AND SMOKE DETECTOR

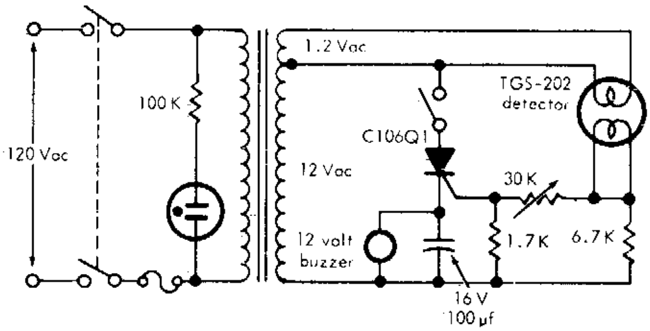


Fig. 40-1

Circuit Notes

This circuit can detect smoke and a number of gases (CO, CO_2, methane, coal gas and others) with a 10 ppm sensitivity. It uses a heated surface semiconductor sensor. Detection occurs when the gas concentration increase causes a decrease of the sensor element internal resistance. The switch in series with the SCR is used for resetting the alarm.

IONIZATION CHAMBER SMOKE DETECTOR

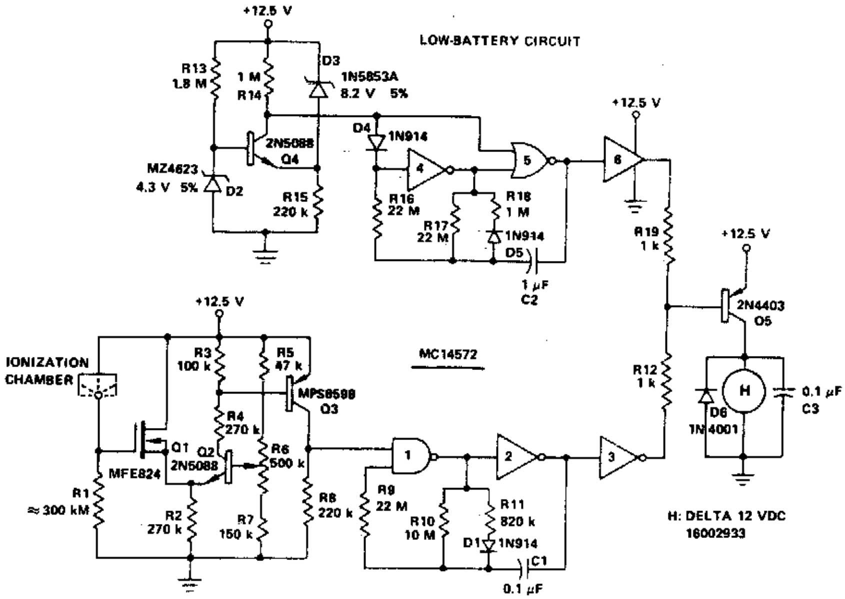


Fig. 40-2

Circuit Notes

Battery-operated, ionization chamber smoke detector includes a circuit to generate a unique alarm when the battery reaches the end of its useful life. The circuit uses the MCMOS MC14572 for two alarm oscillators (smoke and low battery). This circuit additionally uses five discrete transistors as buffers and comparators.

IONIZATION CHAMBER SMOKE DETECTOR

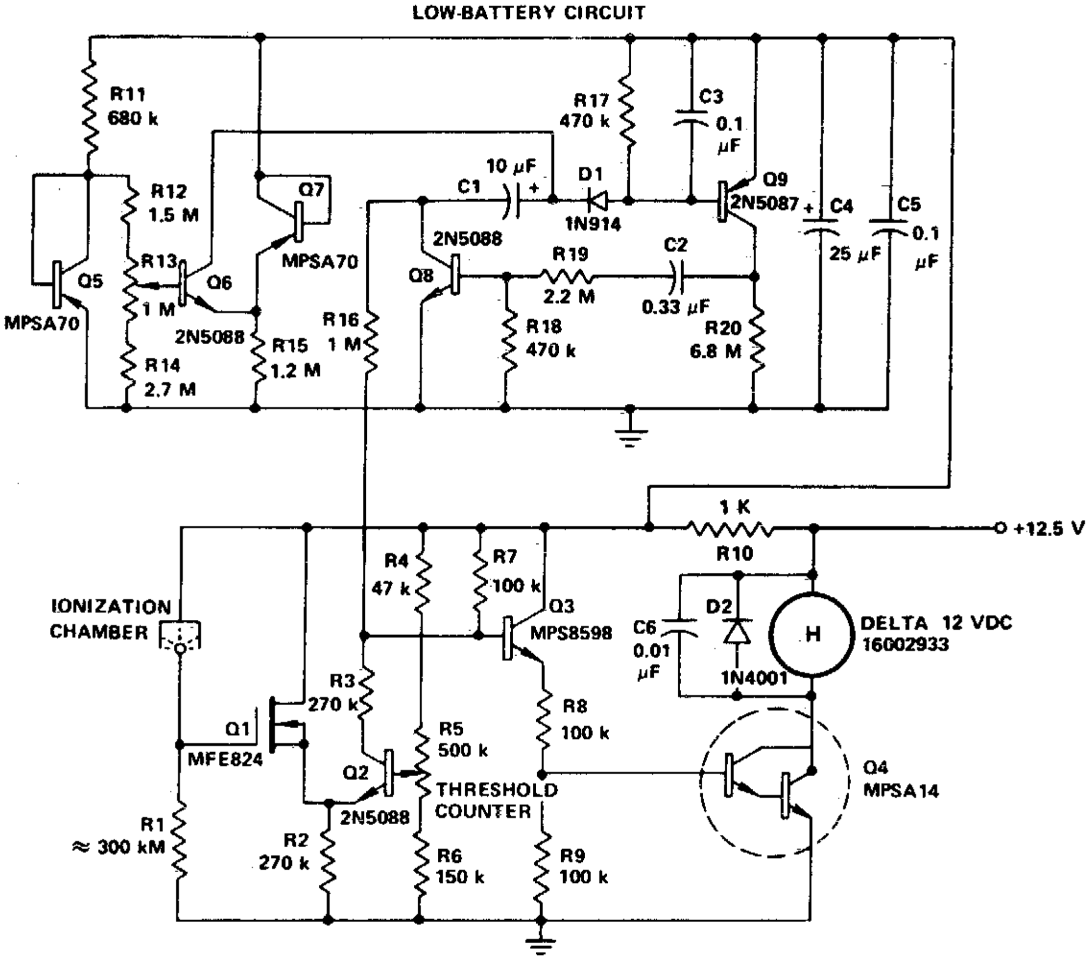


Fig. 40-3

Circuit Notes

If the smoke alarm signal must be a continuous one rather than pulsating, then the slightly less expensive, all discrete transistor version of the MC14572 may be used.

41

Indicators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ten-Step Voltage-Level Indicator
Beat Frequency Indicator
Three-Step Level Indicator
Indicator and Alarm
Five-Step Voltage-Level Indicator
Visible Voltage Indicator
Voltage Level Detector
Zero Center Indicator for FM Receivers
Visual Zero-Beat Indicator

TEN-STEP VOLTAGE-LEVEL INDICATOR

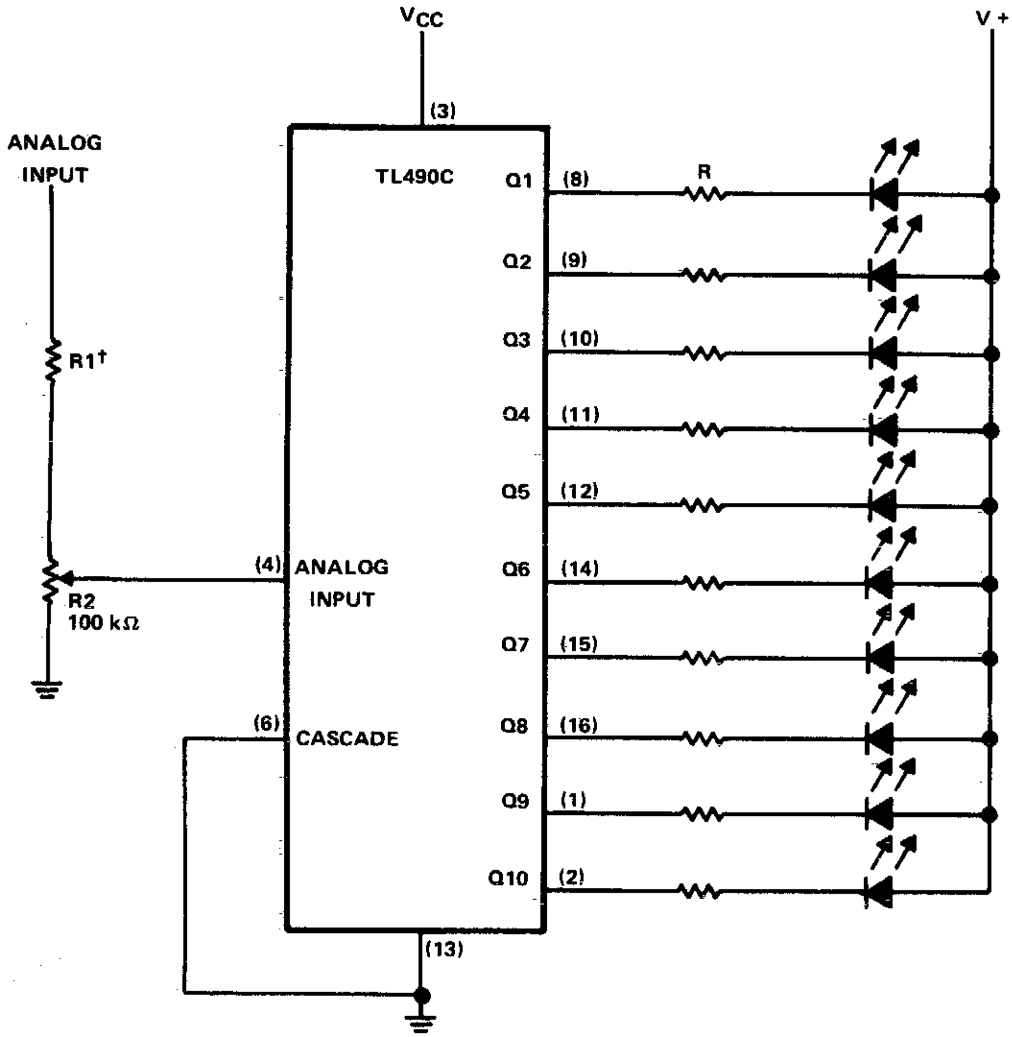


Fig. 41-1

Circuit Notes

This ten-step adjustable analog level detector is capable of sinking up to 40 milliamperes at each output. The voltage range at the input pin should range from 0 to 2 volts. Circuits of this type are useful as liquid-level indicators, pressure indicators, and temperature indicators. They may also be used with a set of active filters to provide a visual indication of harmonic content of audio signals.

BEAT FREQUENCY INDICATOR

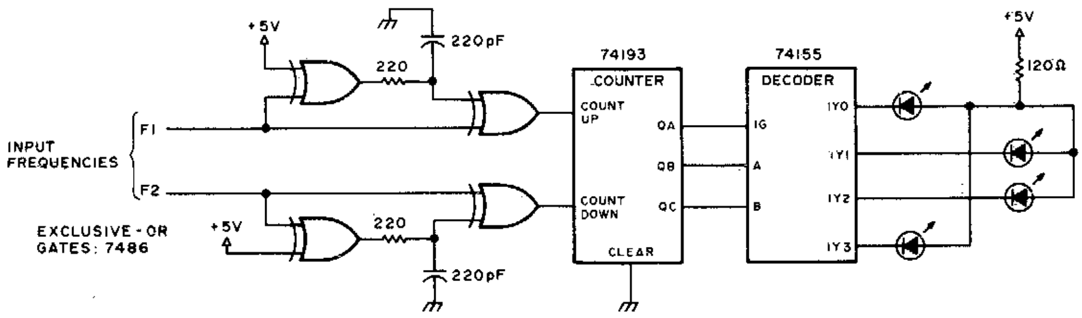


Fig. 41-2

Circuit Notes

This circuit uses LEDs to display the beat frequency of two-tone oscillators. Only one LED is on at a time, and the apparent rotation of the dot is an exact indication of the best frequency. When f1 is greater than f2, a dot of light rotates clockwise; when f1 is less than f2, the dot rotates counterclockwise; and when f1 equals f2, there is no rotation.

THREE-STEP LEVEL INDICATOR

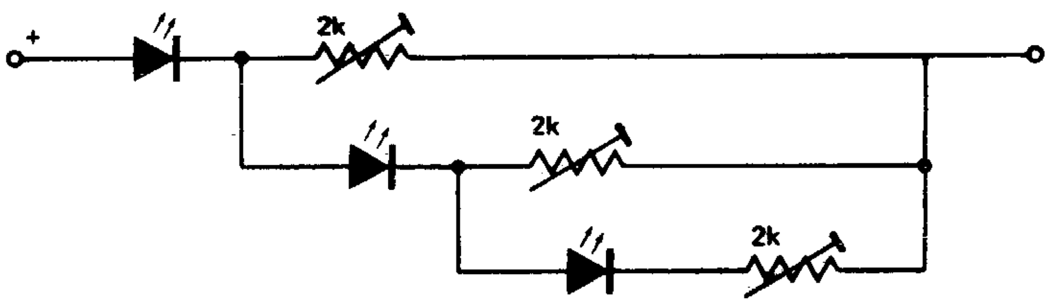
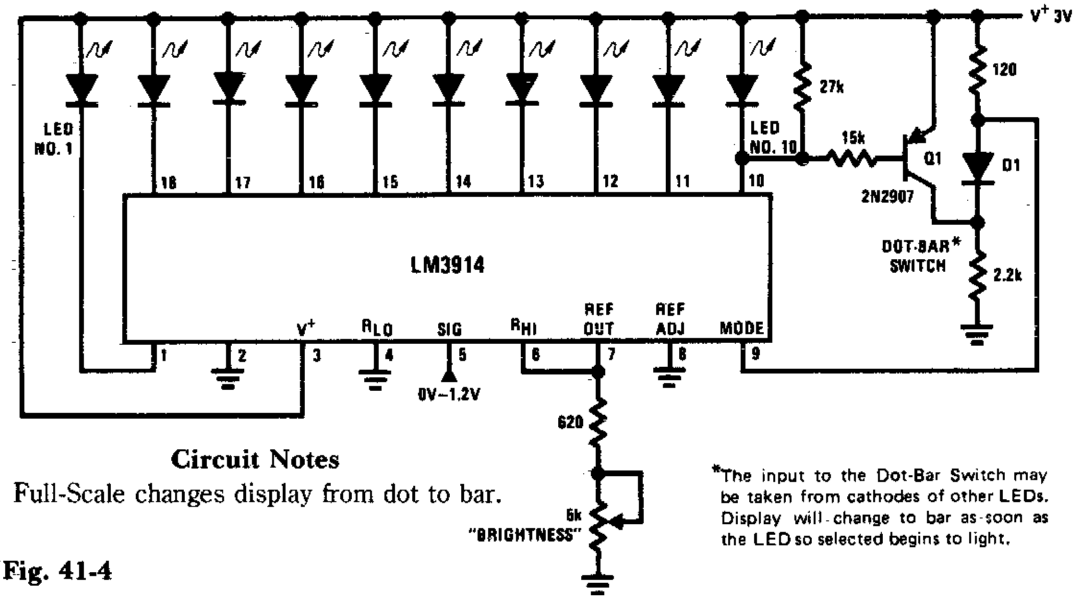


Fig. 41-3

Circuit Notes

This circuit makes a very compact level indicator where a meter would be impractical or not justified due to cost. Resistor values will depend on type of LED used. For MV50 LEDs the resistors are 2 K for steps of approx 2 V and current drain with all three LEDs on of 5 mA. The chain can be extended but current drain increases rapidly and the first LED carries all the current drawn from the supply.

INDICATOR AND ALARM



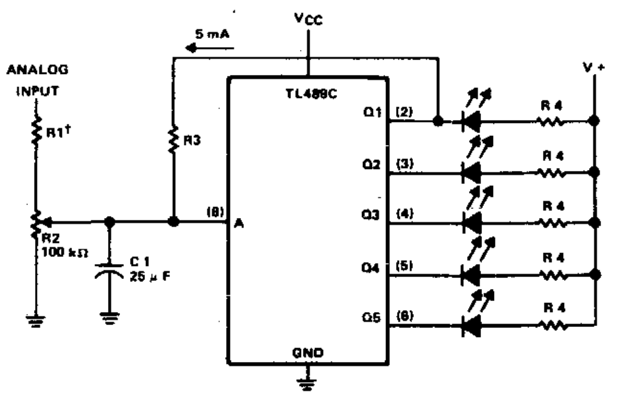
Circuit Notes

Full-Scale changes display from dot to bar.

*The input to the Dot-Bar Switch may be taken from cathodes of other LEDs. Display will change to bar as soon as the LED so selected begins to light.

Fig. 41-4

FIVE-STEP VOLTAGE-LEVEL INDICATOR



† R1 is chosen to ensure that the voltage across R2 is less than 8 volts. Normally it will be set to 1 volt.

Fig. 41-5

Circuit Notes

This circuit provides a visual indication of the input analog voltage level. It has a high input impedance at pin 8 and open-collector outputs capable of sinking up to 40 milliamperes. It is suitable for driving a linear array of 5 LEDs to indicate the level is 5 steps. The voltage at the analog input should be in the range of zero to approximately one volt and should never exceed eight volts.

VISIBLE VOLTAGE INDICATOR

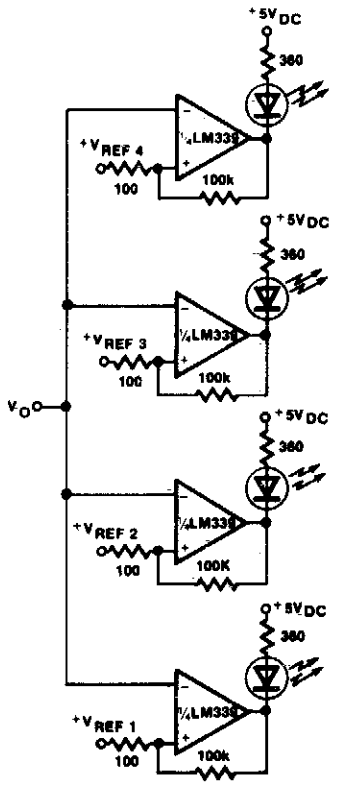


Fig. 41-6

VOLTAGE LEVEL DETECTOR

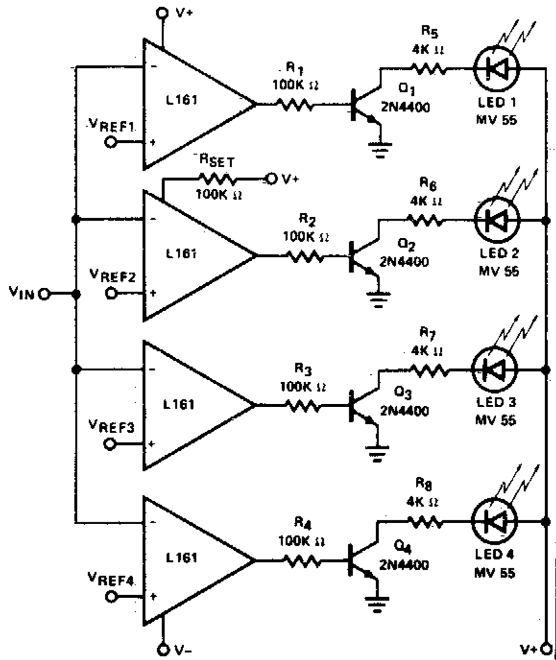


Fig. 41-8

ZERO CENTER INDICATOR FOR FM RECEIVERS

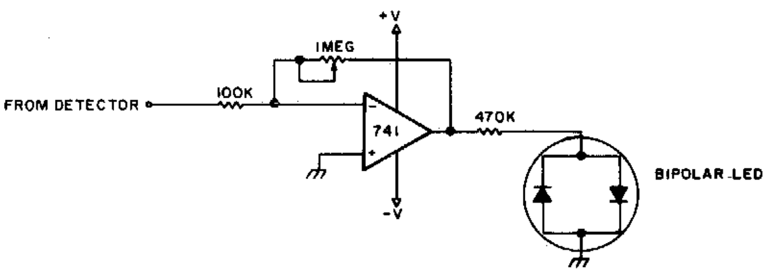
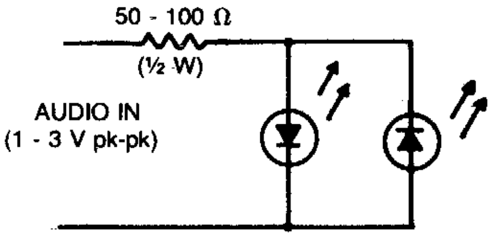


Fig. 41-7

Circuit Notes

To adjust, tune in a station and adjust the 1 M pot for a null. Then ask the station to modulate and fine adjust so modulation peaks don't light the LEDs. Stations are properly tuned when neither LED is lit.

VISUAL ZERO-BEAT INDICATOR



LEDs: FAIRCHILD FLV-100 RED,
OR MONSANTO MV-5094 RED/RED,
OR MONSANTO MV-5491 RED/GREEN

Fig. 41-9

Circuit Notes

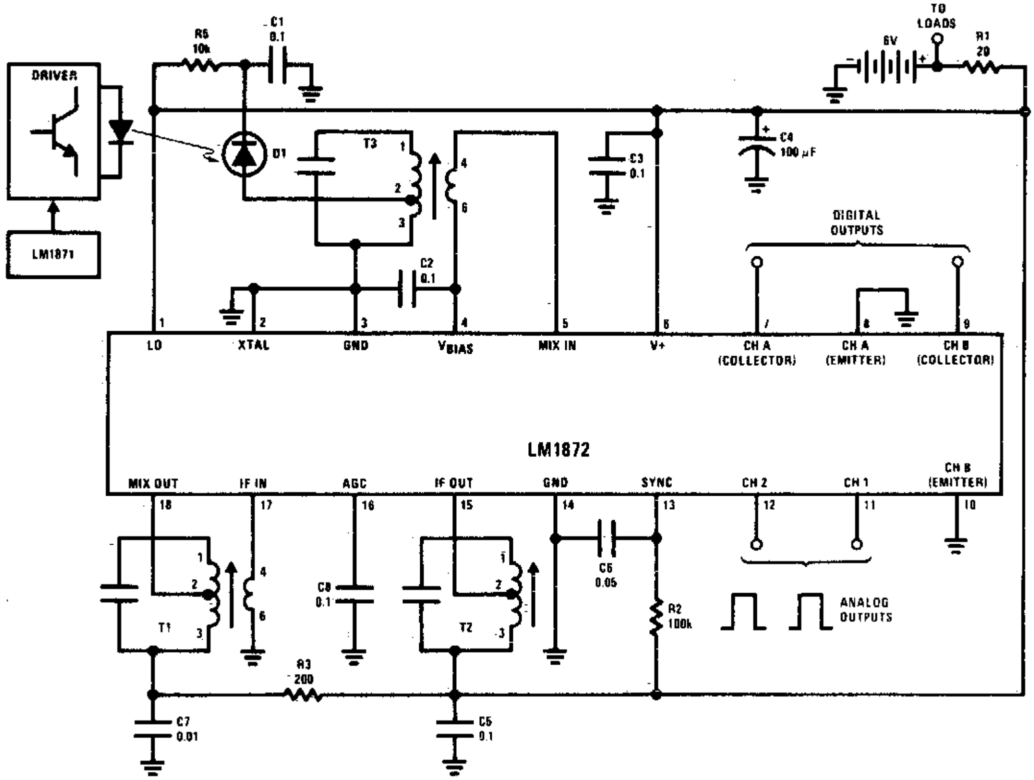
Light-emitting diodes connected with reverse polarity provide a visual indication of zero-beat frequency. Each LED is on for only half a cycle of the input. When the input frequency is more than 1 kilohertz away from the zero-beat frequency, both LEDs appear to be on all the time. As the input frequency comes within about 20 hertz of zero beat, the LEDs will flicker until zero beat is reached. Both LEDs glow or flicker until zero beat is reached, when they go out.

42

Infrared Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

IR TYPE DATA LINK



R1 - Load decoupling

R2 - Sync timer; $R2 = \frac{t}{0.7\ C6}$, $R2 \leq 470k$

R3 - Preamp decoupling

R5 - Photodiode decoupling

C1 - Photodiode decoupling

C2 - V_{BIAS} bypass

C3 - V^+ bypass

C4 - Load decoupling

C5 - IF bypass; optional

C6 - Sync timer; $C6 = \frac{t_{SYNC}}{0.7\ R2}$, $C6 \leq 0.5\ \mu F$

C7 - Preamp decoupling

C8 - AGC

T1 - 455 kHz preamp transformer
Toko* 10 EZC type (RMC-502182), Qu = 110
Pin 1-2, 82T; pin 2-3, 82T
Pin 1-3, 164T; pin 4-6, 30T

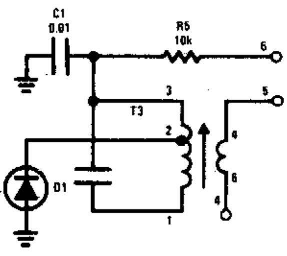
T2 - 455 kHz IF transformer
Toko* 10 EZC type (RMC-402503), Qu = 110
Pin 1-2, 98T; pin 2-3, 66T
Pin 1-3, 164T; pin 4-6, 8T

T3 - 455 kHz input transformer
Toko* 10 EZC type (RMC-202313), Qu = 110
Pin 1-2, 131T; pin 2-3, 33T
Pin 1-3, 164T; pin 4-6, 5T

D1 - PN or PIN Silicon Photodiode

	Photodiode, D1	Active Area (cm^2)
Vactec	VTS 5088	0.18
Vactec	VTS 6089	0.52
UDT	PIN 5D or 6 DP	0.20
UDT	PIN 220 DP	2.0
Siemens	BPY 12	0.20

* Toko America, Inc.
5520 West Touhy Ave.
Skokie, Ill. 60077
(312)677-3640 Tlx: 72-4372



Input Stage Where the Case of D1 Is Connected to the Anode

Fig. 42-1

IR REMOTE CONTROL TRANSMITTER/RECEIVER

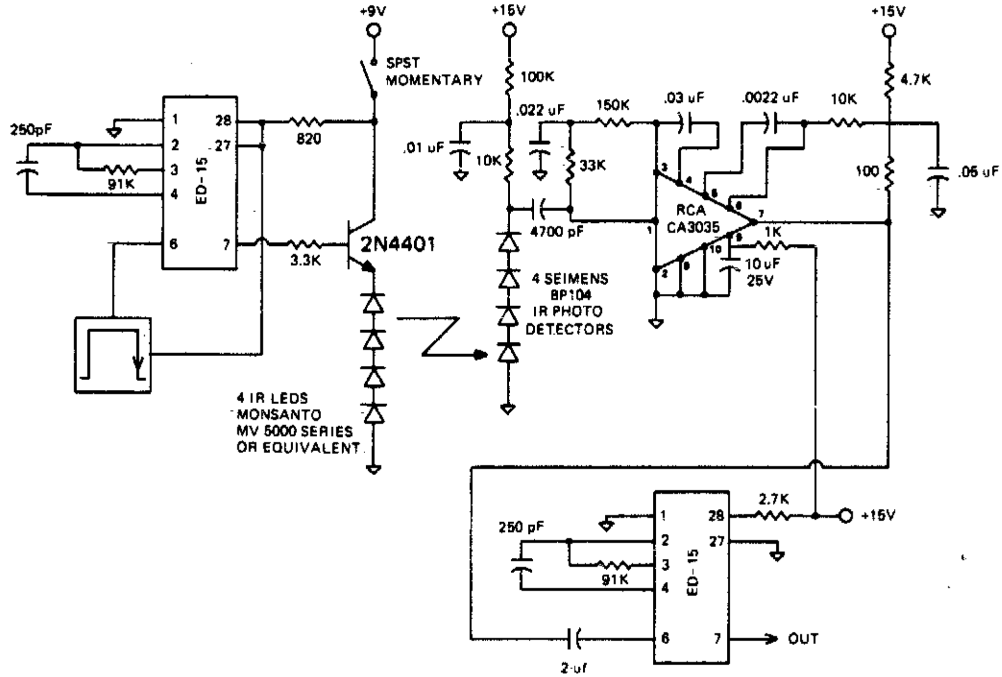


Fig. 42-2

Circuit Notes

The circuit is designed to operate at 25 kHz. The data stream turns the 2N4401 hard on or off depending upon the coded state. This in turn switches the series infrared LEDs on and off. The receiver circuit consists of a three stage amplifier with photo diodes arrayed for maximum coverage of the reception area. The range of this set-up should be about 10 meters.

COMPACT IR RECEIVER

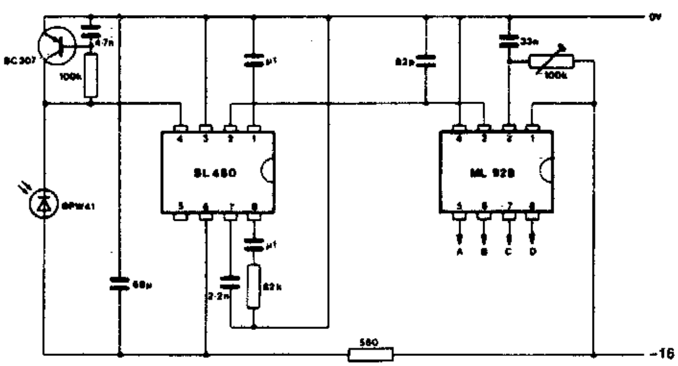


Fig. 42-3

IR TRANSMITTER

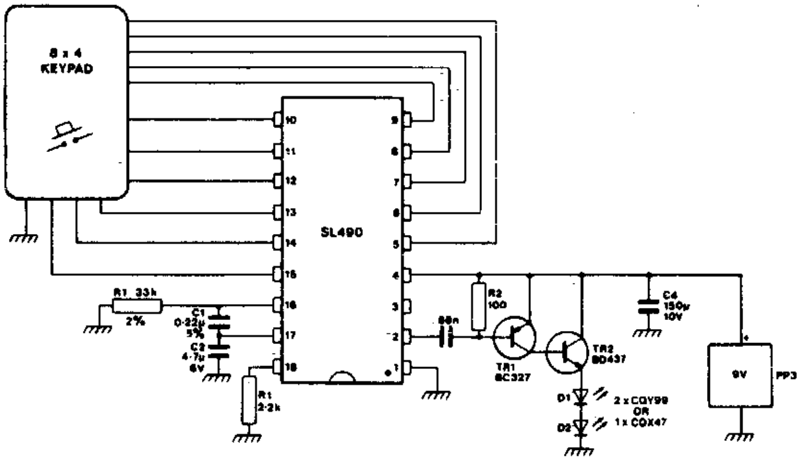


Fig. 42-4

Circuit Notes

This simple infra-red transmitter, where the PPM output from pin 2 of the SL490 is fed to the base of the PNP trasmitter TR1, produces an amplified current pulse about 15 μsec wide. This pulse is further amplified by TR2 and applied to the infra-red diodes D1 and D2.

REMOTE LOUDSPEAKER VIA IR LINK

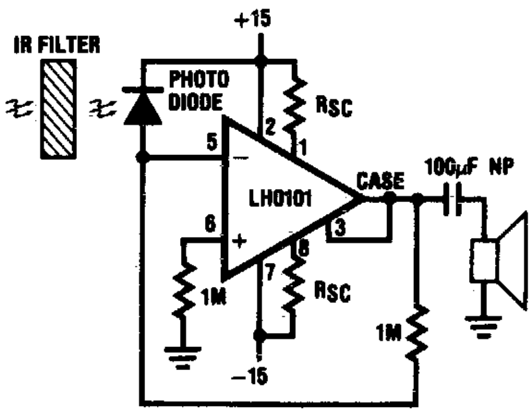


Fig. 42-5

PROXIMITY DETECTOR

NOTE:
IC1 IS CA3240
Q1 IS 2N3819
Q2,4 ARE BC184L
Q3 IS BD140

D1 IS PHOTODIODE
D2 IS 1N4148
ZD1 IS 2V7 400mW ZENER
LED1 IS 3mm RED LED
LED2 IS IS INFRA-RED LED

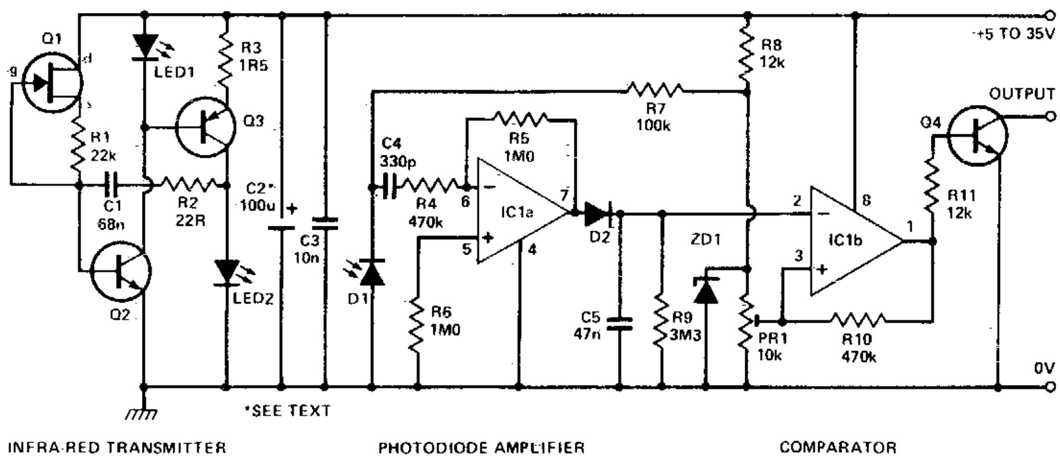


Fig. 42-6

Circuit Notes

This circuit provides a means of detecting the presence of anything by the reflection of infra-red light and provides a direct digital output of object detection. By the use of modulation and high power bursts of infra-red at a very low duty cycle, a detection range of over a foot is achieved. Works on the principle of transmitting a beam of modulated infra-red light from the emitter diode LED2, and receiving reflections from objects passing in front of the beam with a photodiode detector D1. The circuit consists of an infra-red transmitter, photodiode amplifier, and a variable threshold comparator.

43

Instrumentation Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

INSTRUMENTATION AMPLIFIER

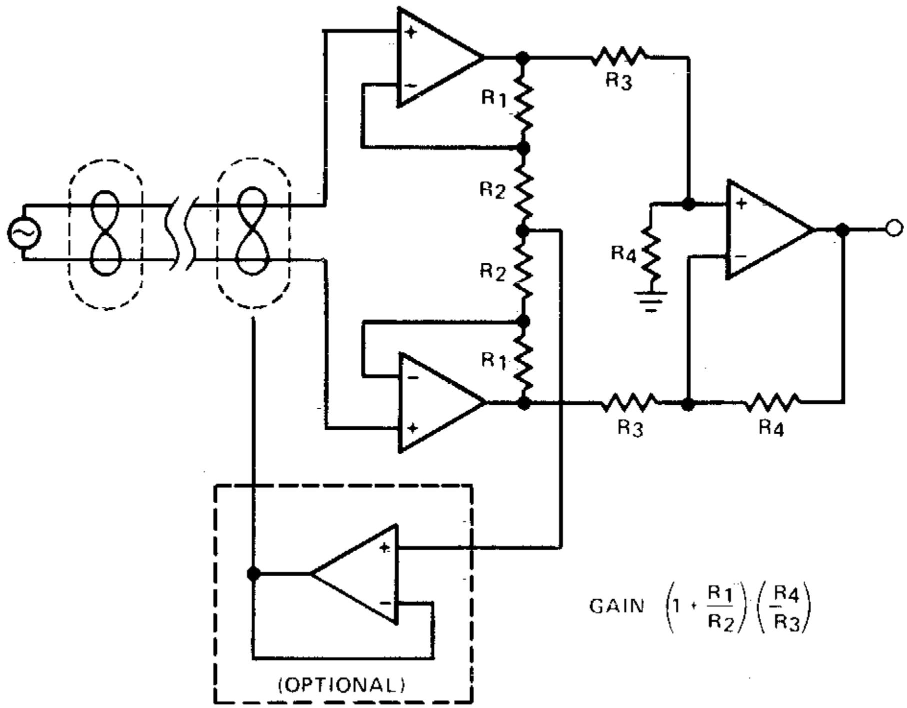


Fig. 43-1

Circuit Notes

Instrumentation amplifiers (differential amplifiers) are specifically designed to extract and amplify small differential signals from much larger common mode voltages. To serve as building blocks in instrumentation amplifiers, op amps must have very low offset voltage drift, high gain and wide bandwidth. The HA-4620/5604 is suited for this application. The optional circuitry makes use of the fourth amplifier section as a shield driver which enhances the ac common mode rejection by nullifying the effects of capacitance-to-ground mismatch between input conductors.

TRIPLE OP-AMP INSTRUMENTATION AMPLIFIER

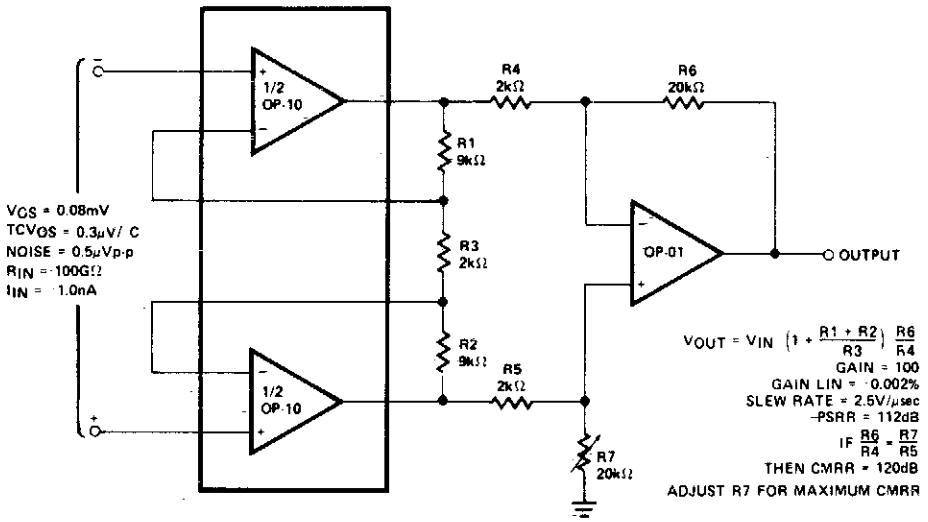


Fig. 43-2

DIFFERENTIAL INPUT INSTRUMENTATION AMPLIFIER WITH HIGH COMMON MODE REJECTION

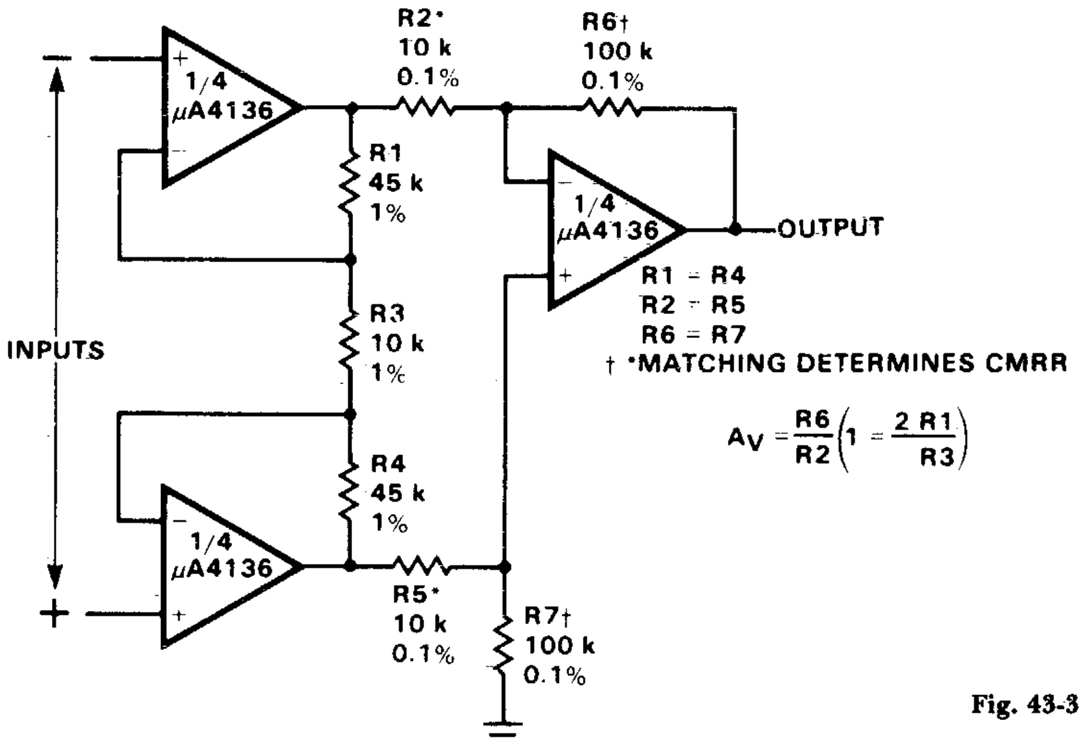
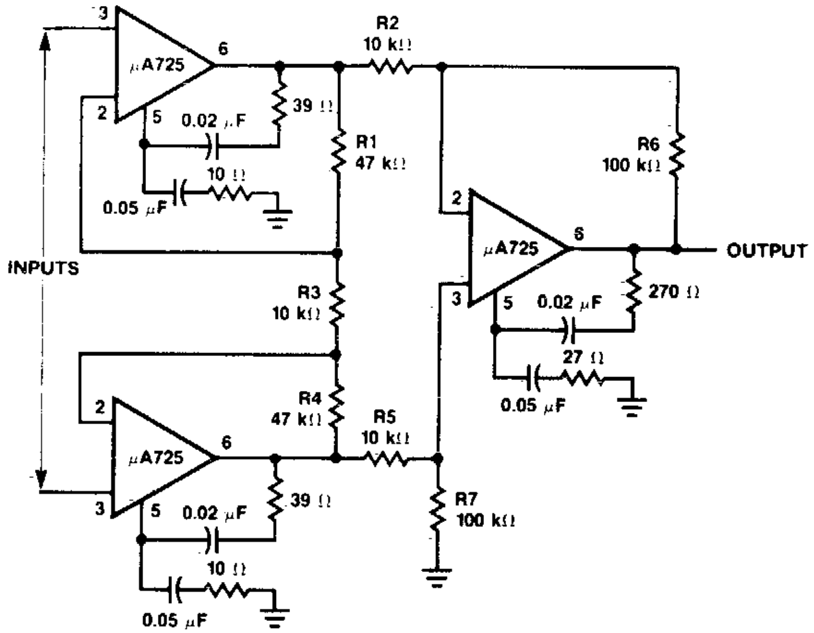


Fig. 43-3

INSTRUMENTATION AMPLIFIER WITH HIGH COMMON MODE REJECTION



$\frac{R1}{R6} = \frac{R3}{R4}$ for best CMRR

$R3 = R4$

$R1 = R6 = 10\ R3$

$\text{Gain} = \frac{R6}{R7}$

Fig. 43-4

LEVEL-SHIFTING ISOLATION AMPLIFIER

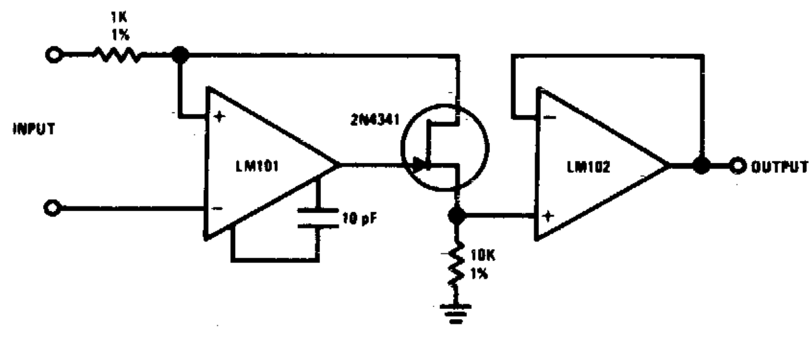


Fig. 43-5

Circuit Notes

The 2N4341 JFET is used as a level shifter between two op amps operated at different power supply voltages. The JFET is ideally suited for this type of application because $I_D = I_S$.

VARIABLE GAIN, DIFFERENTIAL-INPUT INSTRUMENTATION AMPLIFIER

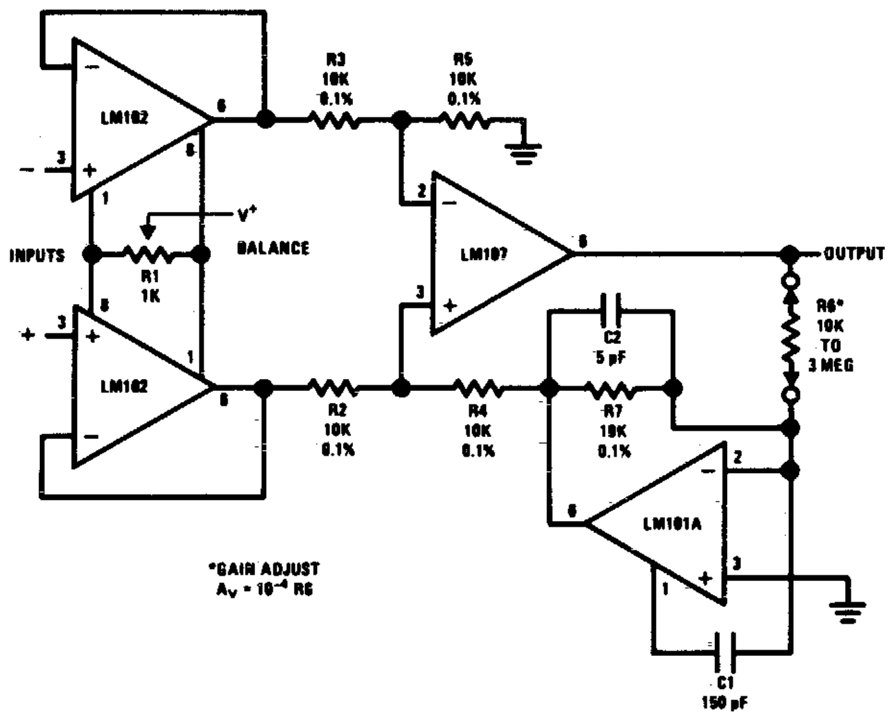


Fig. 43-6

INSTRUMENTATION AMPLIFIER

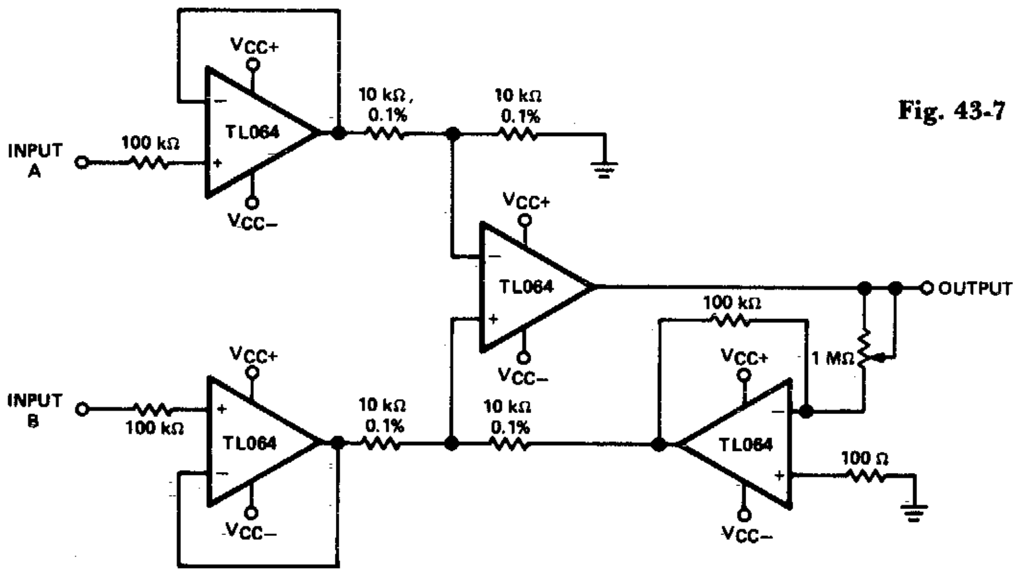


Fig. 43-7

LOW SIGNAL LEVEL, HIGH IMPEDANCE INSTRUMENTATION AMPLIFIER

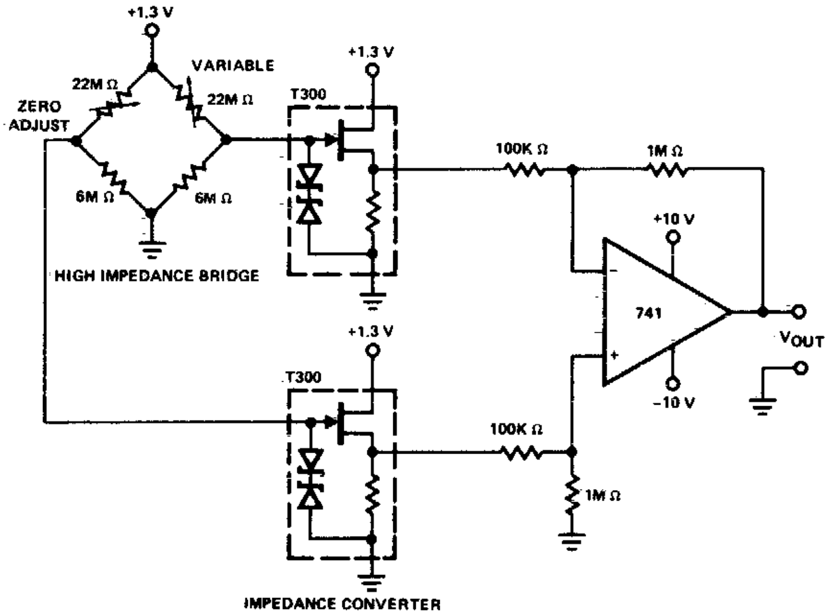


Fig. 43-8

CHOPPER CHANNEL AMPLIFIER

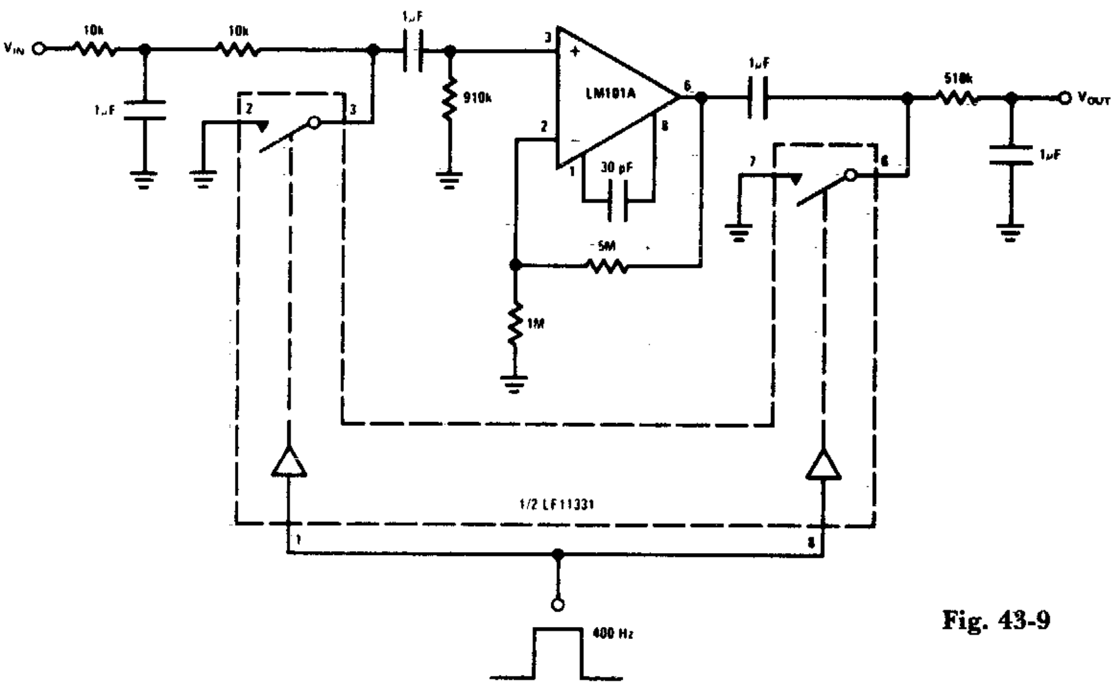
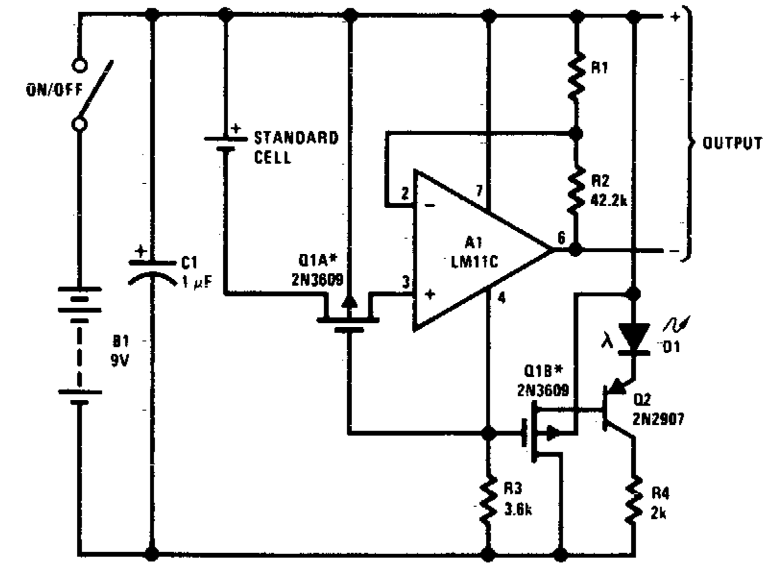


Fig. 43-9

BATTERY POWERED BUFFER AMPLIFIER FOR STANDARD CELL



* cannot have gate protection diode; $V_{TH} > V_{OUT}$

Fig. 43-10

Circuit Notes

This circuit has negligible loading and disconnects the cell for low supply voltage or overload on output. The indicator diode extinguishes as disconnect circuitry is activated.

BRIDGE TRANSDUCER AMPLIFIER

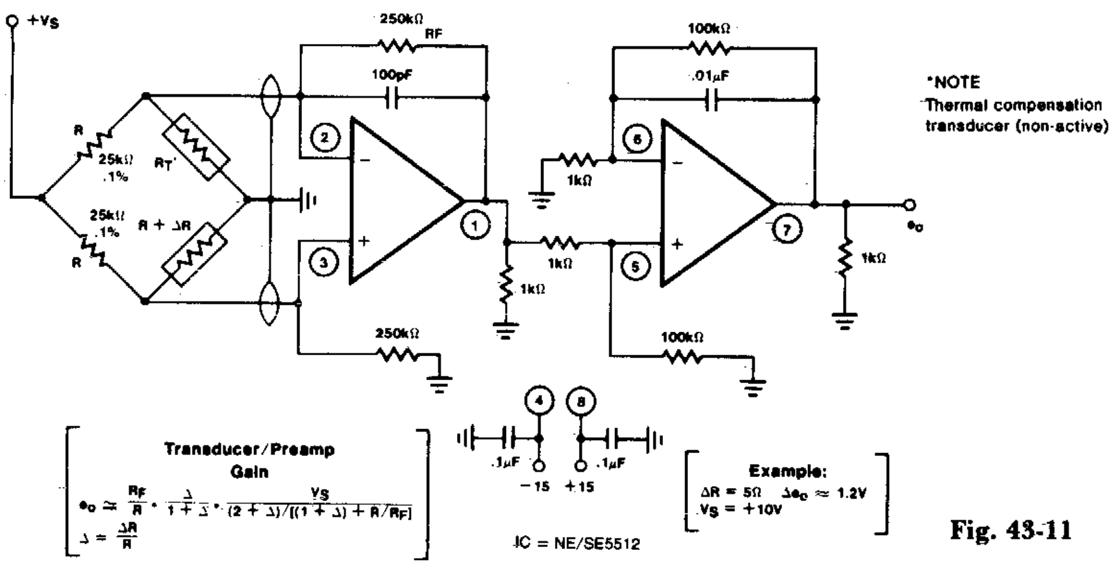


Fig. 43-11

INSTRUMENTATION AMPLIFIER

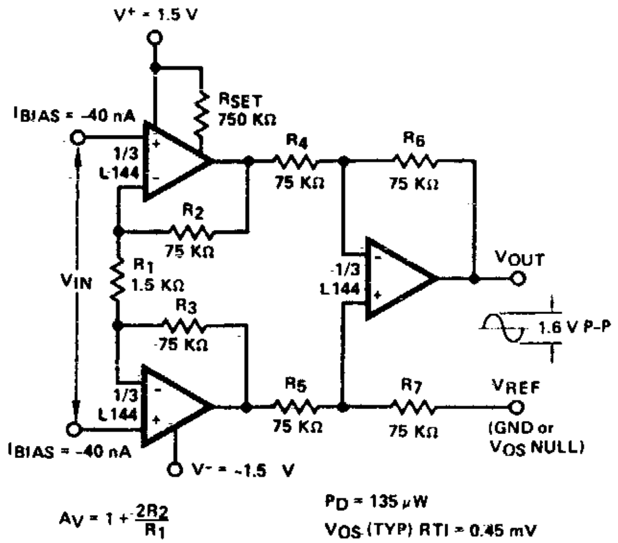


Fig. 43-12

Circuit Notes

Three-amplifier circuit consumes only 135 μW of power from a $\pm$1.5 V power supply. With a gain of 101, the instrumentation amplifier is ideal in sensor interface and biomedical preamplifier applications. The first stage provides all of the gain while the second stage is used to provide common mode rejection and double-ended to single-ended conversion.

ISOLATION AMPLIFIER FOR MEDICAL TELEMETRY

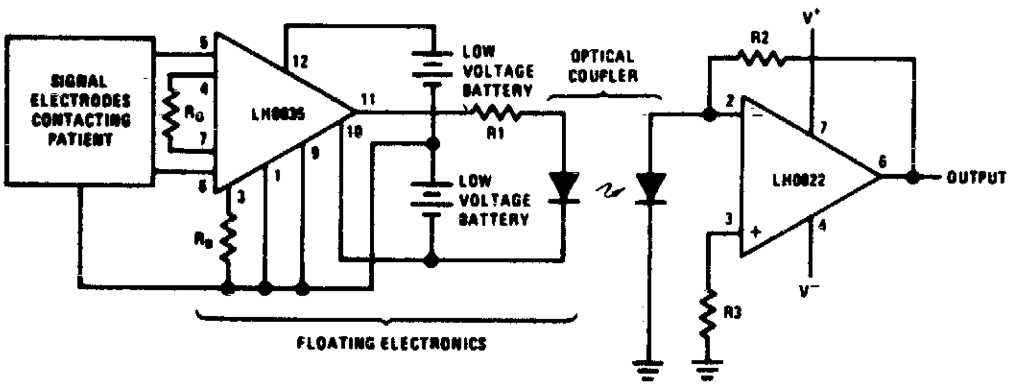


Fig. 43-13

HIGH GAIN DIFFERENTIAL INSTRUMENTATION AMPLIFIER

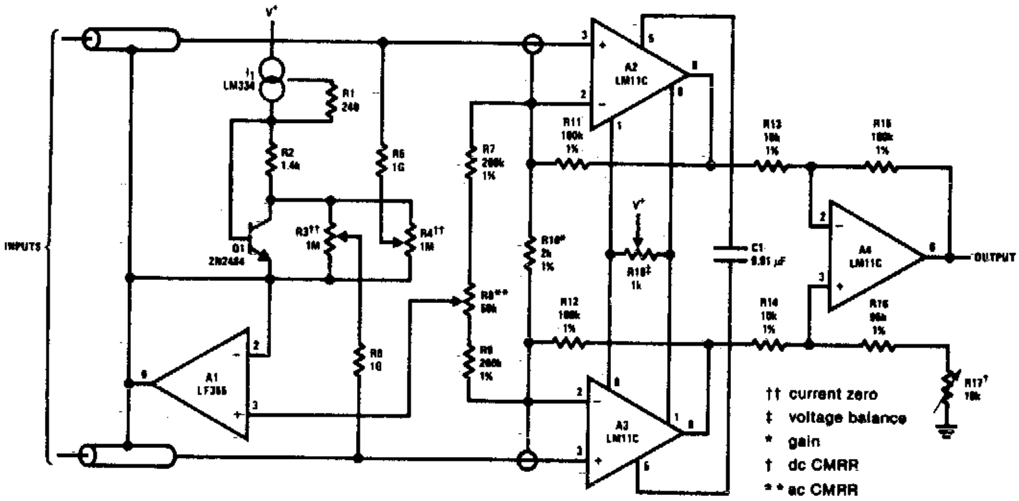


Fig. 43-14

Circuit Notes

This circuit includes input guarding, cable bootstrapping, and bias current compensation. Differential bandwidth is reduced by C1 which also makes common-mode rejection less dependent on matching of input amplifiers.

HIGH IMPEDANCE BRIDGE AMPLIFIER

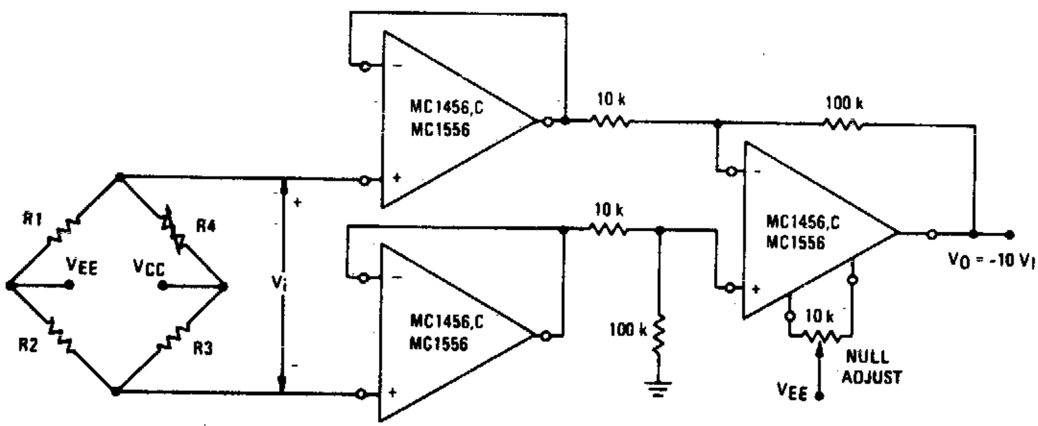


Fig. 43-15

INSTRUMENTATION AMPLIFIER (TWO OP AMP DESIGN)

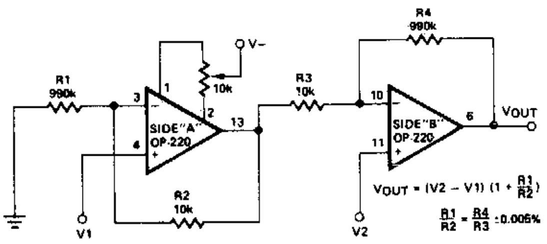


Fig. 43-16

HIGH IMPEDANCE DIFFERENTIAL AMPLIFIER

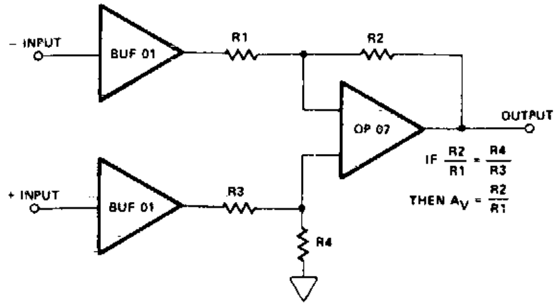


Fig. 43-19

INSTRUMENTATION AMPLIFIER

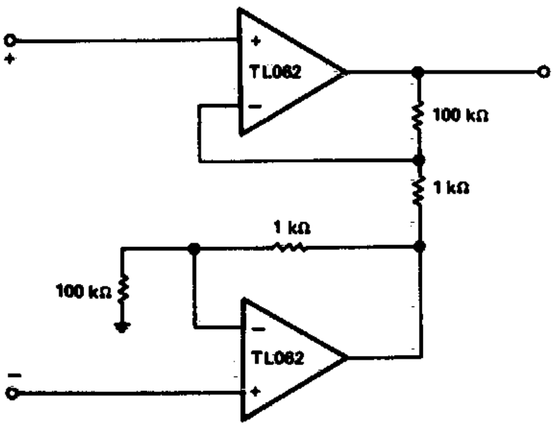
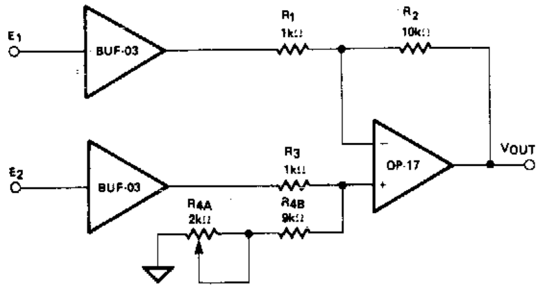


Fig. 43-17

HIGH SPEED INSTRUMENTATION AMPLIFIER



NOTE: $\frac{R_2}{R_1} = \frac{R_4}{R_3}$ FOR GOOD COMMON MODE REJECTION.
R_{4A} IS ADJUSTED FOR BEST CMRR.

Fig. 43-20

DIFFERENTIAL INPUT INSTRUMENTATION AMPLIFIER

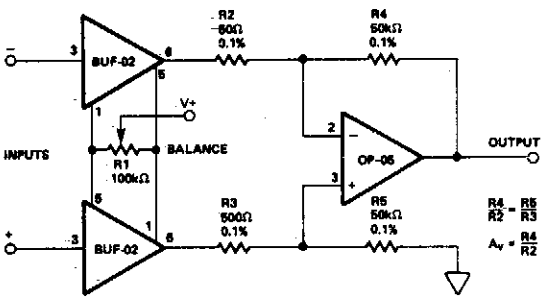
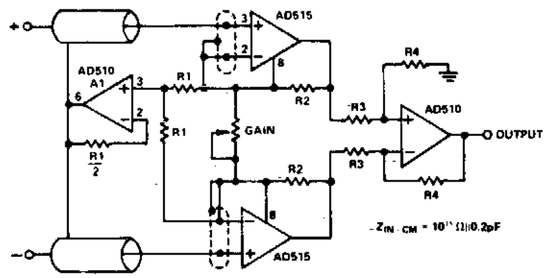


Fig. 43-18

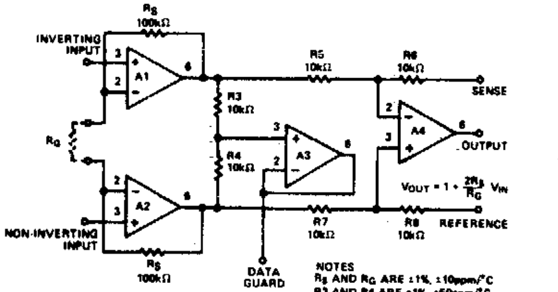
VERY HIGH IMPEDANCE INSTRUMENTATION AMPLIFIER



(ALL RESISTORS OF SAME NUMBER SHOULD BE MATCHED ±0.1%)
(BUFFER A1 BOOSTS COMMON MODE Z_{IN} BY DRIVING CABLE SHIELDS AT COMMON MODE VOLTAGE AND NEUTRALIZING CM CAPACITANCE)

Fig. 43-21

PRECISION FET INPUT INSTRUMENTATION AMPLIFIER



NOTES
R_S AND R_G ARE ±1%, ±10ppm/°C
R3 AND R4 ARE ±1%, ±50ppm/°C.
R5, R6, R7, R8 ARE A MATCHED NETWORK, ±0.01%, ±2ppm/°C TRACKING TC.

ICs are AD547L.

Fig. 43-22

HIGH STABILITY THERMOCOUPLE AMPLIFIER

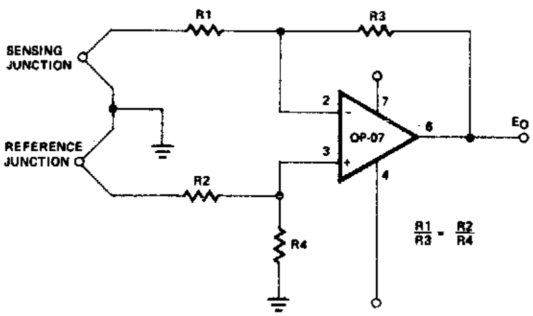


Fig. 43-23

HIGH STABILITY THERMOCOUPLE AMPLIFIER

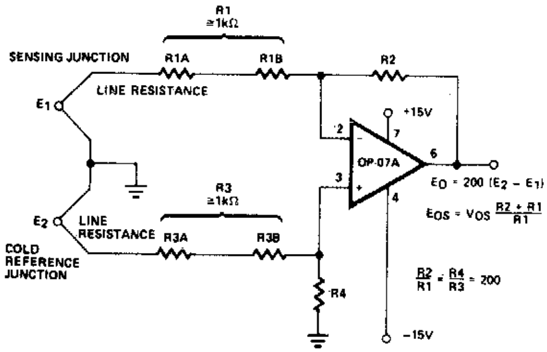
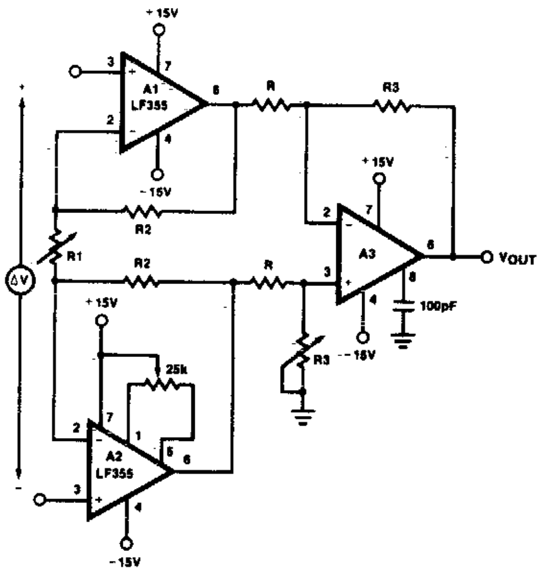


Fig. 43-24

HIGH IMPEDANCE LOW DRIFT INSTRUMENTATION AMPLIFIER



- $V_{OUT} = \frac{R3}{R}\left[\frac{2R2}{R1} + 1\right]\Delta V$, $V- + 2V \le V_{IN}$ Common-Mode $\le V+$
- System V_{OS} adjusted via A2 V_{OS} adjust
- Trim R3 to boost up CMRR to 120dB.

Fig. 43-25

44

Light Activated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PULSE GENERATION BY INTERRUPTING A LIGHT BEAM

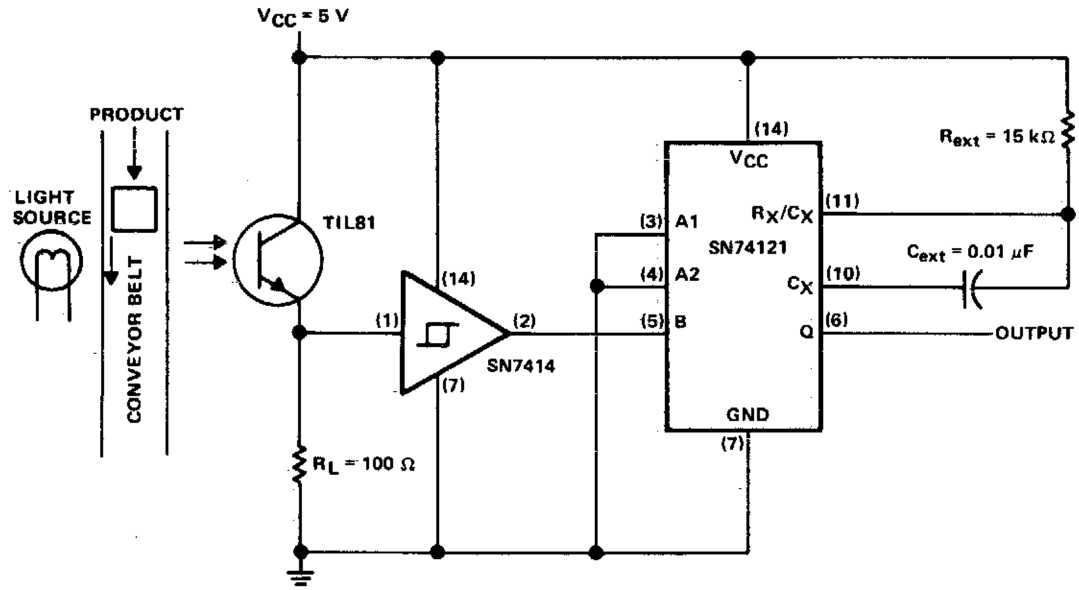
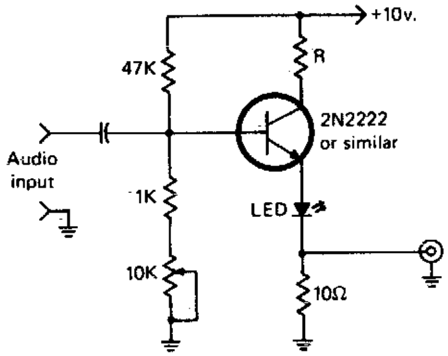


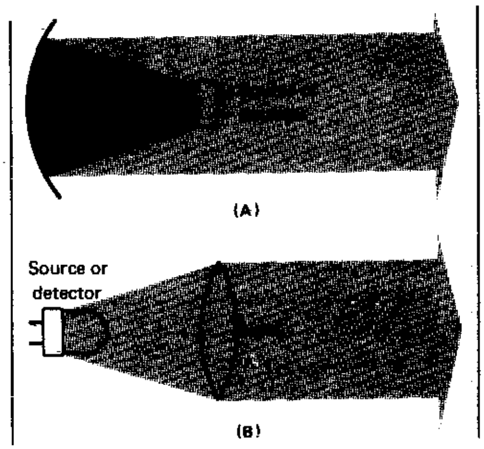
Fig. 44-1

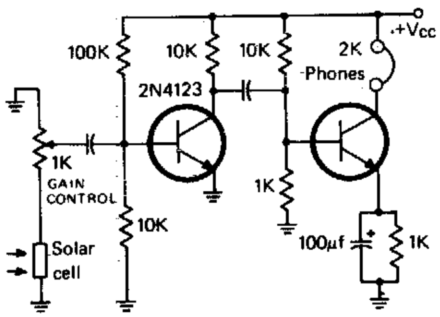
Circuit Notes

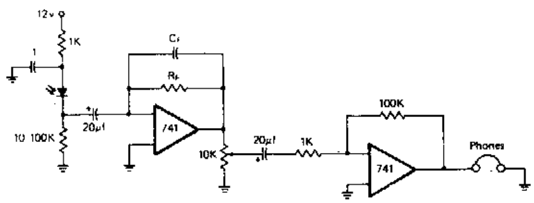
This circuit puts out a pulse when an object on the conveyor belt blocks the light source. The light source keeps the phototransistor turned on. This produces a high-logic-level voltage at the Schmitt-trigger inverter and a TTL-compatible low logic level at pin 5 of the monostable. When an object blocks the light, TIL81 turns off the Schmitt-trigger inverter to triggers the one shot.

OPTICAL COMMUNICATION SYSTEM









Circuit Notes

The simple modulator stage will accommodate most common LEDs. By adjusting the potentiometer, the bias of the transistor is varied until the LED is at its half output point. Then, audio will cause it to vary above and below this point. The purpose of R1 is to limit the current through the LED to a safe level and the purpose of the 10 ohm resistor is to allow a portion of the modulating signal to be observed on a scope.

Fig. 44-2

FOUR QUADRANT PHOTO-CONDUCTIVE DETECTOR AMPLIFIER

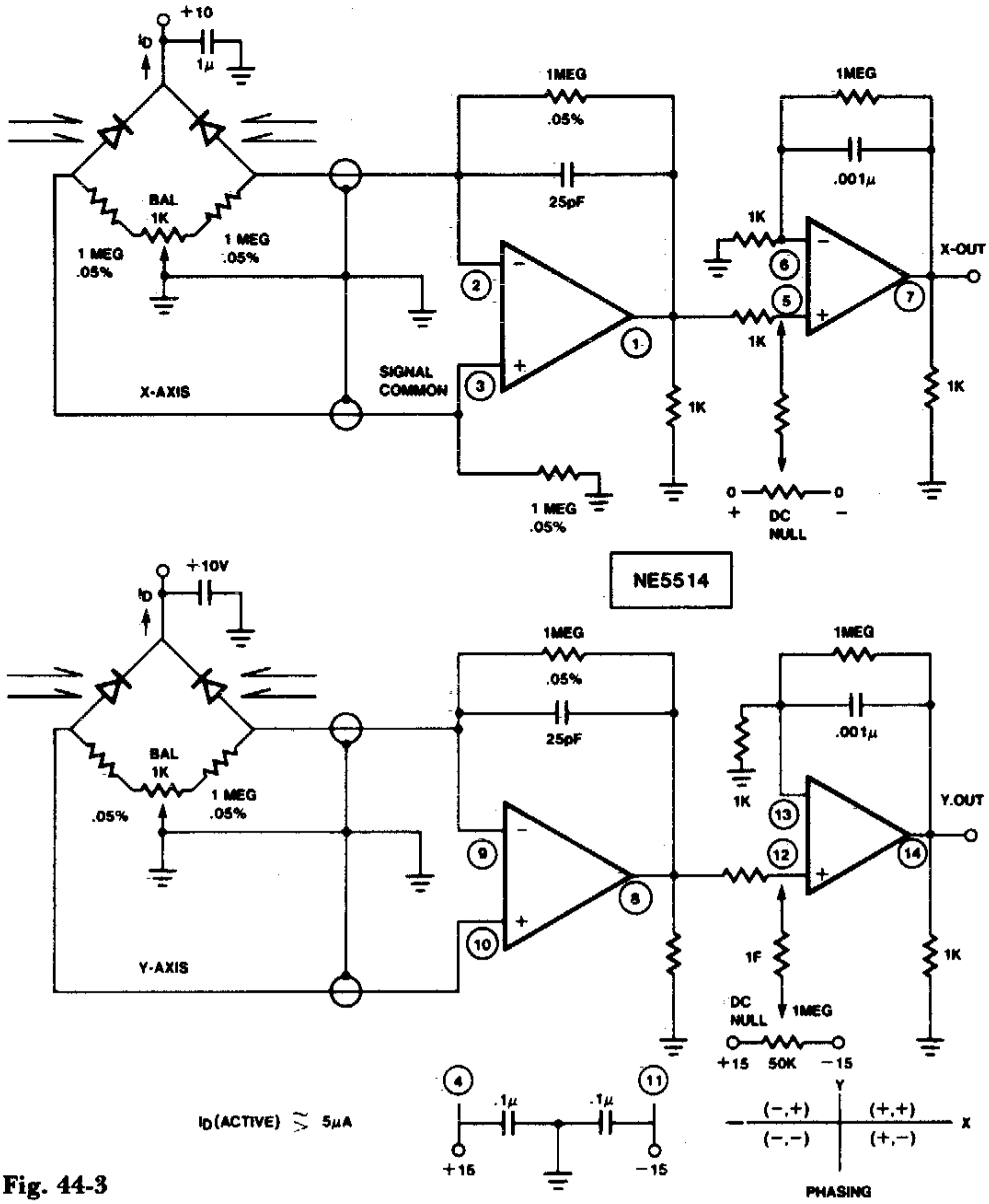


Fig. 44-3

Circuit Notes

Use this circuit to sense four quadrant motion of a light source. By proper summing of the signals from the X and Y axes, four quadrant output may be fed to an X-Y plotter, oscilloscope, or computer for simulation. IC = NE/SE5514

PRECISION PHOTODIODE COMPARATOR

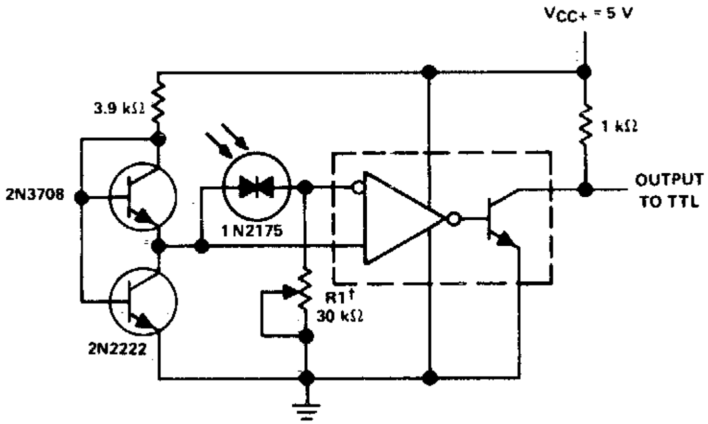


Fig. 44-4

Circuit Notes

R1 sets the comparison level. At comparison, the photodiode has less than 5 mV across it, decreasing dark current by an order of magnitude. IC = LM 111/211/311.

AUTOMATIC NIGHT LIGHT

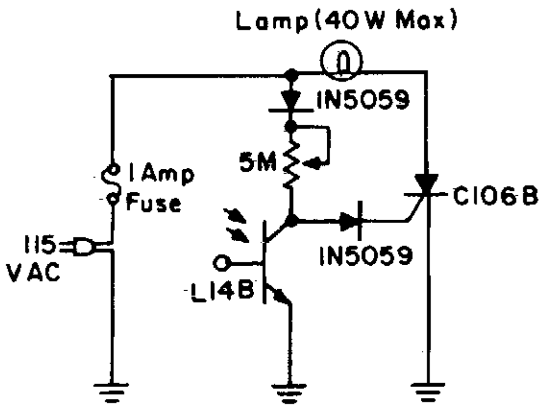


Fig. 44-5

Circuit Notes

During daylight hours, the L14B photo-Darlington (JEDEC registered as 2N5777 through 2N5780) shunts all gate current to ground. At night, the L14B effectively provides a high resistance, diverting the current into the gate of the C106B and turning on the lamp.

RECEIVER FOR 50 kHz FM OPTICAL TRANSMITTER

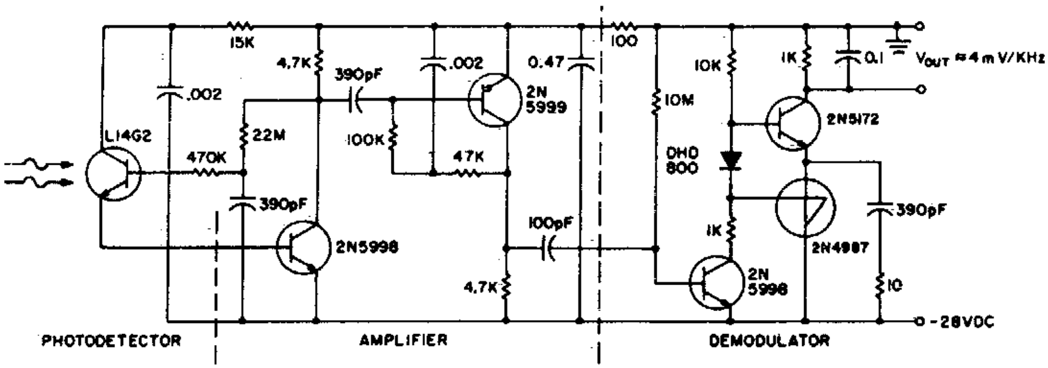


Fig. 44-6

Circuit Notes

This circuit consists of a L14G2 detector, two stages of gain, and a FM demodulator. Better sensitivity can be obtained using more stages of stabilized gain with AGC.

PHOTODIODE AMPLIFIER

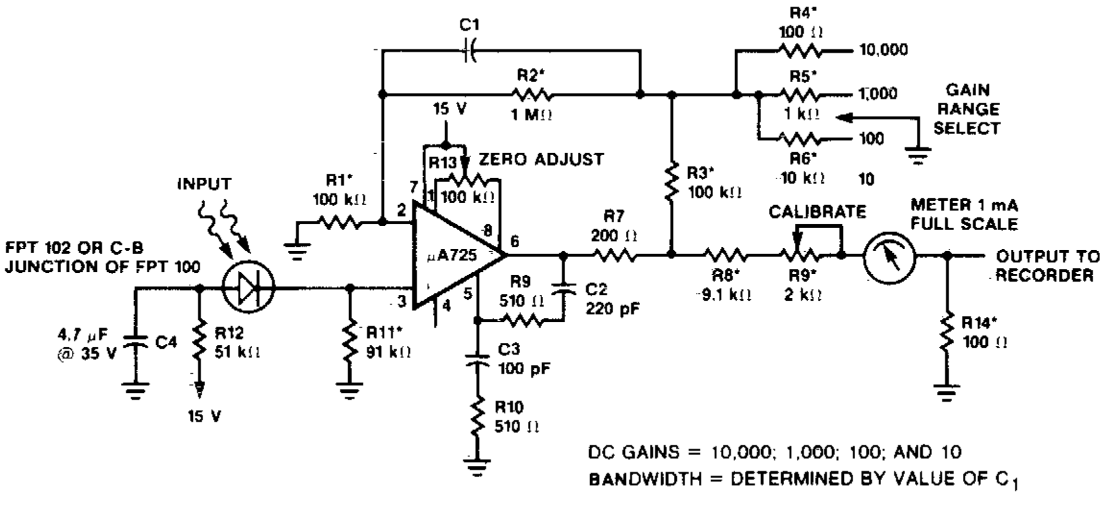


Fig. 44-7

OPTICAL SCHMITT TRIGGER

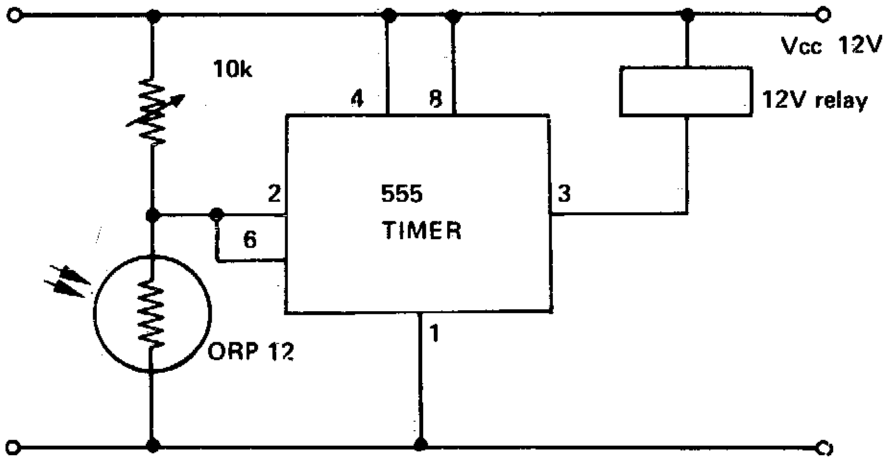


Fig. 44-8

Circuit Notes

This circuit shows a 555 with its trigger and threshold inputs connected together used to energize a relay when the light level on a photoconductive cell falls below a preset value. Circuit can be used in other applications where a high input impedance and low output impedance are required with the minimum component count.

ADJUSTABLE LIGHT DETECTION SWITCH

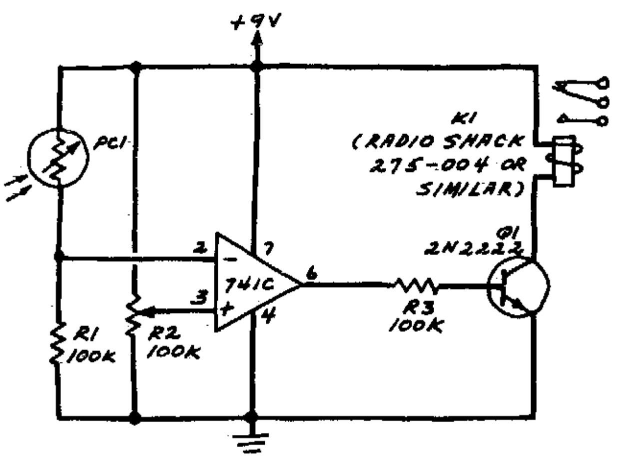


Fig. 44-9

Circuit Notes

R2 sets the circuit's threshold. When the light intensity at PCI's surface is decreased, the resistance of PC1 a cadmium-sulfide photoresistor is increased. This decreases the voltage at the inverting input of the 741. When the reference voltage at the 741's noninverting input is properly adjusted via R2, the comparator will switch from low to high when PC1 is darkened. This turns on Q1 which, in turn, pulls in relay K1.

PHOTOCELL MEMORY SWITCH FOR AC POWER CONTROL

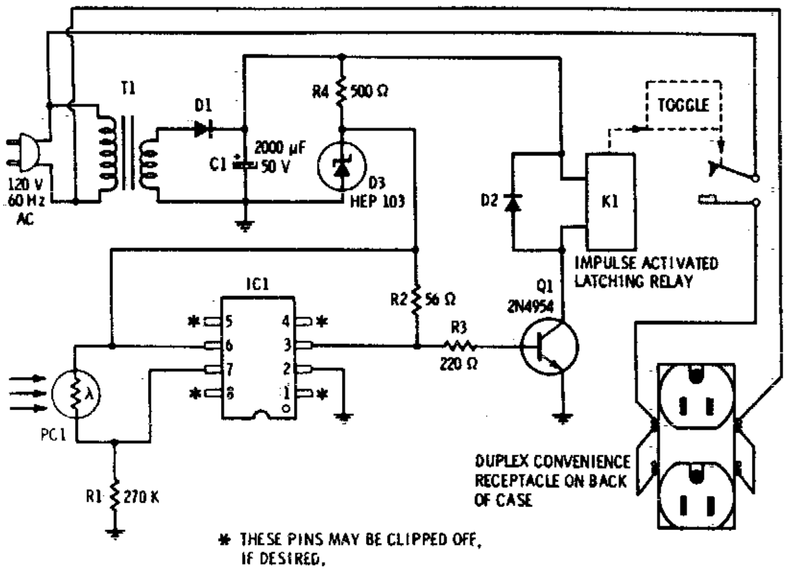
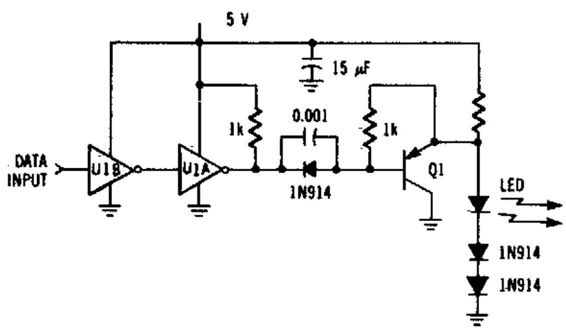


Fig. 44-10

Circuit Notes

Provides remote control for ac-powered devices by using the beam of a flashlight as a magic wand. The important aspect of this gadget is that it remembers. Activate it once to apply power to a device and it stays on. Activate it a second time and power goes off and stays off. It consists of a combination of a high-sensitivity photocell, a high-gain IC Schmitt trigger, and an impulse-actuated latching relay.

OPTICAL TRANSMITTER



Circuit Notes

Driver circuit uses an MC74LS04 and one discrete transistor. The circuit can drive the LED (MFOE1200) at up to 1 Mbps data rate.

Fig. 44-11

LIGHT INTERRUPTION DETECTOR

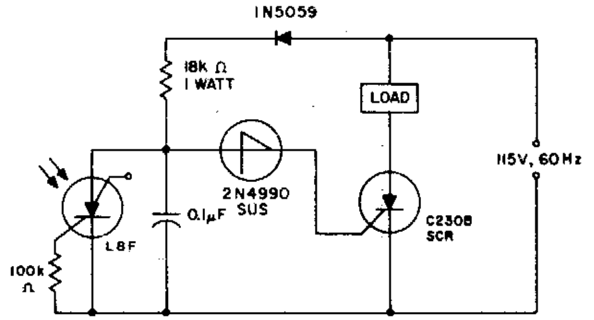
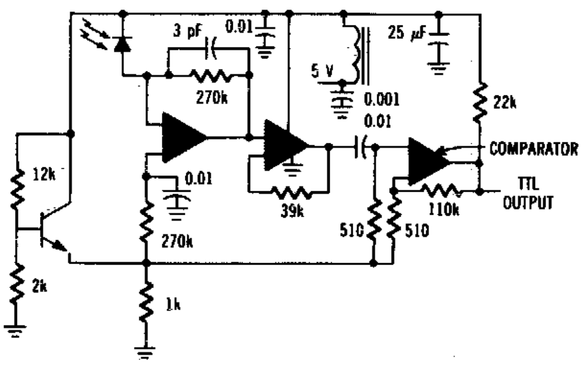


Fig. 44-12

Circuit Notes

When the light incident on the LASCR is interrupted, the voltage at the anode to the 2N4990 unilateral switch goes positive on the next positive cycle of the power which in turn triggers the switch and the C230 SCR when the switching voltage of the unilateral switch is reached. This will cause the load to be energized for as long as light is not incident on the LASCR.

OPTICAL RECEIVER



Circuit Notes

The MFOD1100 PIN diode requires shielding from emi.

Fig. 44-13

LIGHT ISOLATED SOLID STATE POWER RELAY CIRCUITS

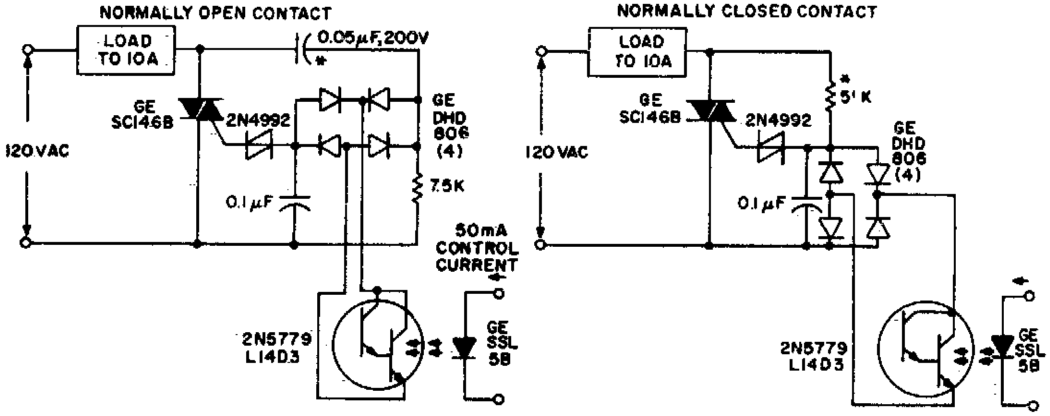


Fig. 44-14

Circuit Notes

Both circuits use the G.E. SC146B, 200 V, 10 A Triac as load current contacts. These triacs are triggered by normal SBS (2N4992) trigger circuits, which are controlled by the photo-Darlington, acting through the DA806 bridge as an ac photo switch. To operate the relays at other line voltages the asterisked (*) components are scaled to supply identical current. Ratings must be changed as required. Incandescent lamps may be used in place of the light emitting diodes, if desired.

PRECISION PHOTODIODE LEVEL DETECTOR

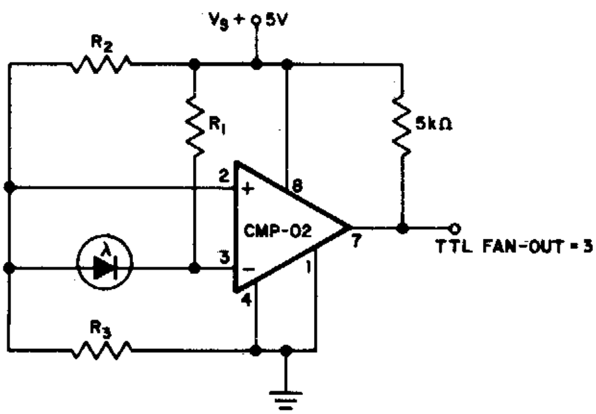
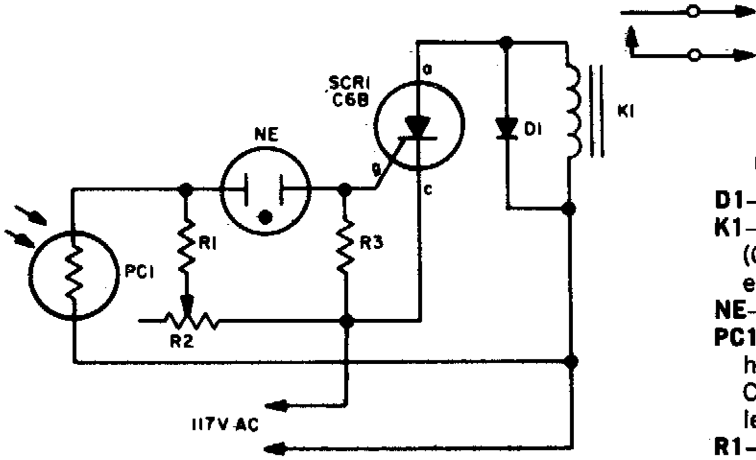


Fig. 44-15

Circuit Notes

For R1 = 2.5 M, R2 = R3 = 5 M. The output state changes at a photo diode current of 0.5 μA.

LIGHT BEAM OPERATED ON-OFF RELAY



PARTS LIST FOR COMMERCIAL KILLER

D1—400-PIV silicon rectifier
K1—117 VAC latching relay (Guardian IR-610L-A115 or equiv.)
NE—NE-83 neon lamp
PC1—Clairex photo cell CL505 for high light level; CL704 or CL705 photocell for low light level
R1—22,000-ohm, ½-watt resistor
R2—1-megohm potentiometer
R3—100-ohm, ½-watt resistor
SCR1—HEP R1218, 200V, 4A, silicon-controlled rectifier

Fig. 44-16

Circuit Notes

When a beam of light strikes the photocell, the voltage across neon lamp NE-1 rises sharply. NE-1 turns on and fires the SCR. K1 is an impulse relay whose contacts stay in position even after coil current is removed. The first impulse opens K1's contacts, the second impulse closes them, etc.

LOGARITHMIC LIGHT SENSOR

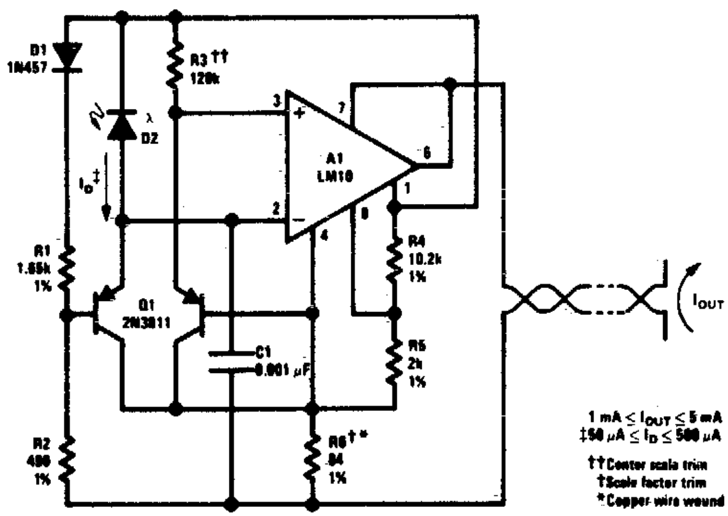


Fig. 44-17

FM (PRM) OPTICAL TRANSMITTER

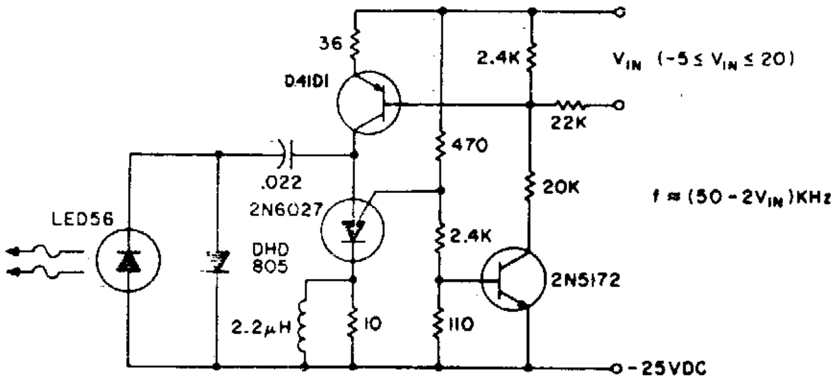
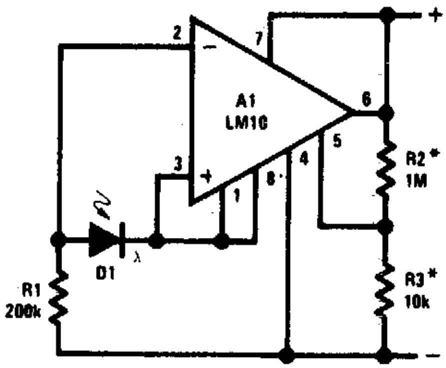


Fig. 44-18

Circuit Notes

The basic circuit can be operated at 80 kHz and is limited by the PUT capacitor combination. 60 kHz is the maximum modulation frequency. The pulse repetition rate is a linear function of V_{IN}, the modulating voltage. Lenses or reflectors minimizes stray light noise effects. Greater output can be obtained by using a larger capacitor, which also gives a lower operating frequency, or using a higher power output IRED such as the F5D1. Average power consumption of the transmitter circuit is less than 3 watts.

LIGHT LEVEL SENSOR



* provides hysteresis

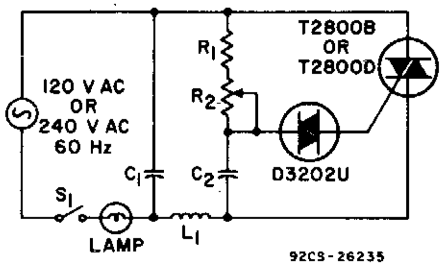
Fig. 44-19

45

Light Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LIGHT DIMMERS



(a) Single-time-constant light-dimmer circuit.

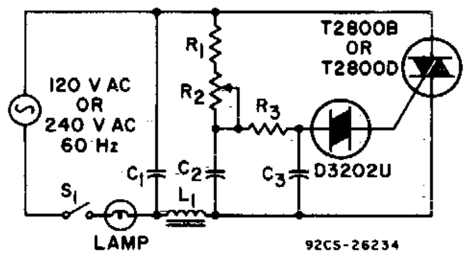
Parts List

120-Volt, 60-Hz Operation

C_1, C_2 = 0.1 μF, 200 V
L_1 = 100 μH
R_1 = 3300 ohms, 0.5 watt
R_2 = light control, potentiometer, 0.25 megohm, 0.5 watt

240-Volt, 50/60 Hz Operation

C_1 = 0.1 μF, 400 V
C_2 = 0.05 μF, 400 V
L_1 = 200 μH
R_1 = 4700 ohms, 0.5 watt
R_2 = light control, potentiometer, 0.25 megohm, 1 watt



(b) Double-time-constant light-dimmer circuit.

Parts List

120-Volt, 60-Hz Operation

C_1, C_2 = 0.1 μF, 200 V
C_3 = 0.1 μF, 100 V
L_1 = 100 μH
R_1 = 1000 ohms, 0.5 watt
R_2 = light control, potentiometer, 0.1 megohm, 0.5 watt

240-Volt, 60-Hz Operation

C_1 = 0.1 μF, 400 V
C_2 = 0.05 μF, 400 V
C_3 = 0.1 μF, 100 V
L_1 = 100 μH
R_1 = 7500 ohms, 2 watts
R_2 = light control, potentiometer, 0.2 megohm, 1 watt
R_3 = 7500 ohms, 2 watts

Fig. 45-1

Circuit Notes

The two lamp-dimmer circuits differ in that (a) employs a single-time-constant trigger network and (b) uses a double-time-constant trigger circuit that reduces hysteresis effects and thereby extends the effective range of the light-control potentiometer. (Hysteresis refers to a difference in the control potentiometer setting at which the lamp turns on and the setting at which the light is extinguished.) The additional capacitor C2 in (b) reduces hysteresis by charging to a higher voltage than capacitor C3. During gate triggering, C3 discharges to form the gate current pulse. Capacitor C2, however, has a longer discharge time constant and this capacitor restores some of the charge removed from C3 by the gate current pulse.

REMOTE CONTROL FOR LAMP OR APPLIANCE

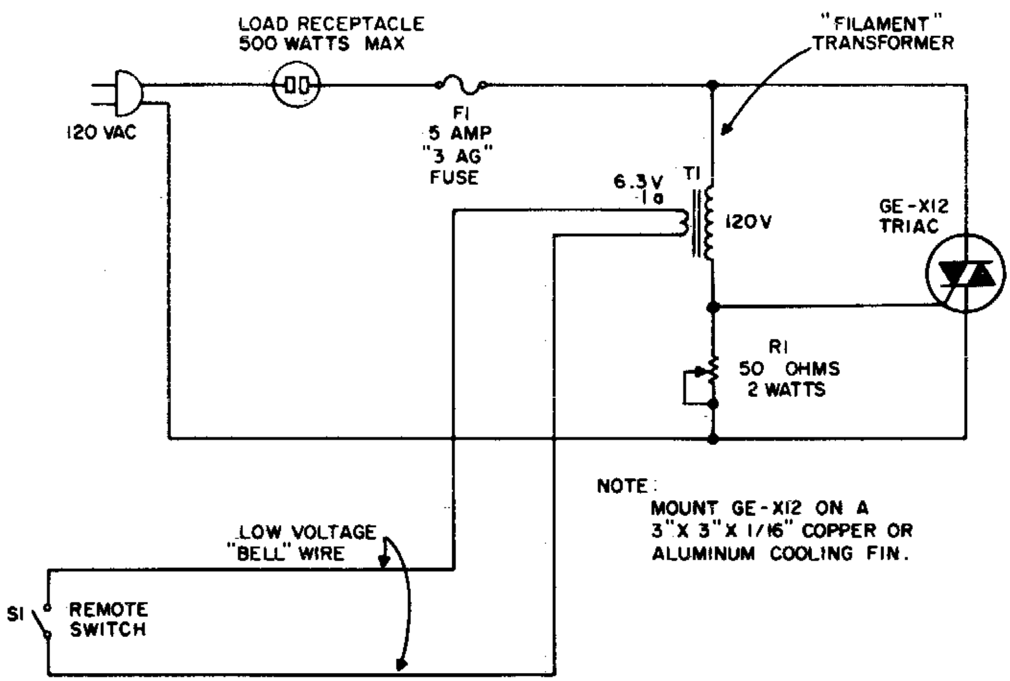


Fig. 45-2

Circuit Notes

The circuit uses the primary current of a small 6.3 volt filament transformer to actuate a triac and energize the load. When switch S1, in the six-volt secondary, of the transformer is open, a small "magnetizing" current flows through the primary winding. This magnetizing current may be large enough to trigger the triac. Therefore, a shunting resistor, R1, is required to prevent such triggering. R1, is adjusted for the highest resistance that will not cause the triac to trigger with S1 open. When single-pole remote switch, S1, closes, the secondary of the transformer is shorted and a high current flows through the 120-volt primary. This triggers the triac and energizes the load. When the triac conducts, current through the primary stops and thus prevents burning out the transformer.

HIGH POWER CONTROL FOR SENSITIVE CONTACTS

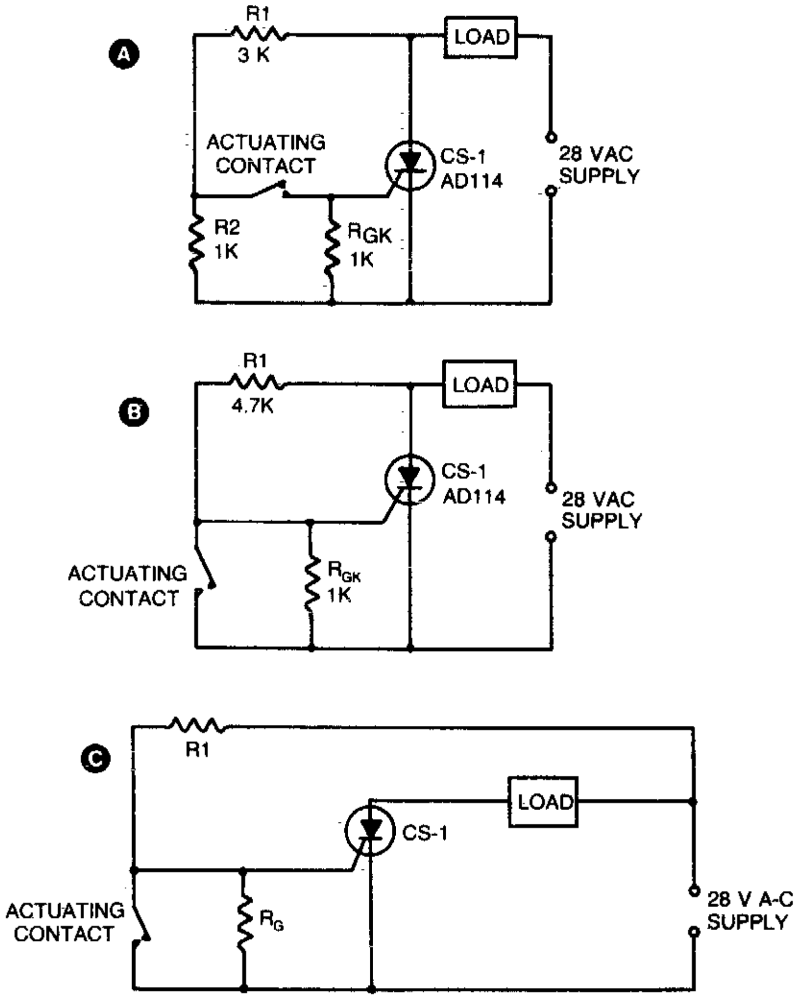


Fig. 45-3

Circuit Notes

Two simple arrangements for resistive loads are shown in A & B. The circuit in A will provide load power when the actuating contact is closed, and no power when the contact is open. B provides the reverse of this action—power being supplied to the load when the contact is open with no load power when the contact is closed. If desired, both circuits can be made to latch by operating with dc instead of the indicated ac supply. In both of these circuits, voltage across the sensitive contacts is under 5 volts, and contact current is below 5 mA. For inductive loads, R1 would normally be returned to the opposite side of the load as shown in C.

COMPLEMENTARY LIGHTING CONTROL

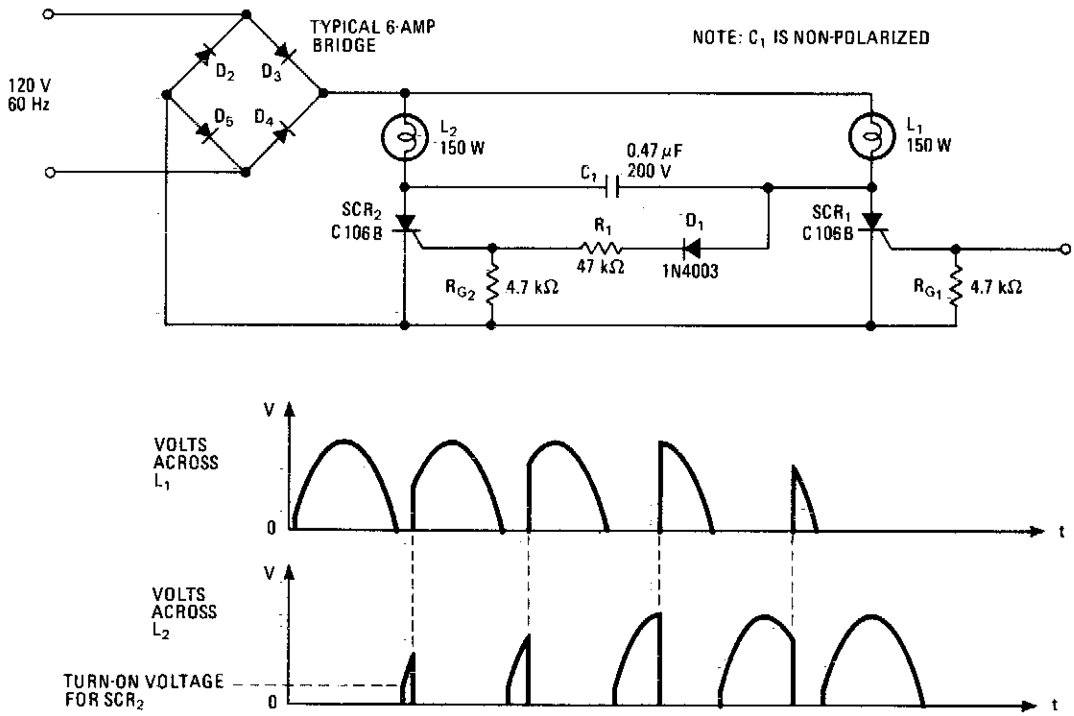


Fig. 45-4

Circuit Notes

This lighting-control unit will fade out one lamp while simultaneously increasing the light output of another. The two loads track each other accurately without adjustments. The gate of SCR1, a silicon-controlled rectifier, is driven from a standard phase-control circuit, based, for example, on a unijunction transistor or a diac. It controls the brightness of lamp L1 directly. Whenever SCR1 is not on, a small current flows through L1, D1, and R1, permitting SCR2 to fire. When SCR1 turns on, current flow ceases through D1 and R1; the energy stored in C1 produces a negative spike that turns SCR2 off.

FLOODLAMP POWER CONTROL

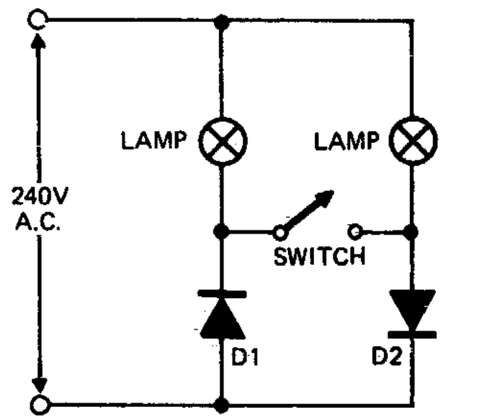


Fig. 45-5

Circuit Notes

When setting up photographic floodlamps, it is sometimes desirable to operate the lamps at lower power levels until actually ready to take the photograph. The circuit allows the lamps to operate on half cycle power when the switch is open, and full power, when the switch is closed. The diodes D1 and D2 should have a 400 volt PIV rating at 5 amps.

HYSTERESIS-FREE PHASE CONTROL CIRCUIT

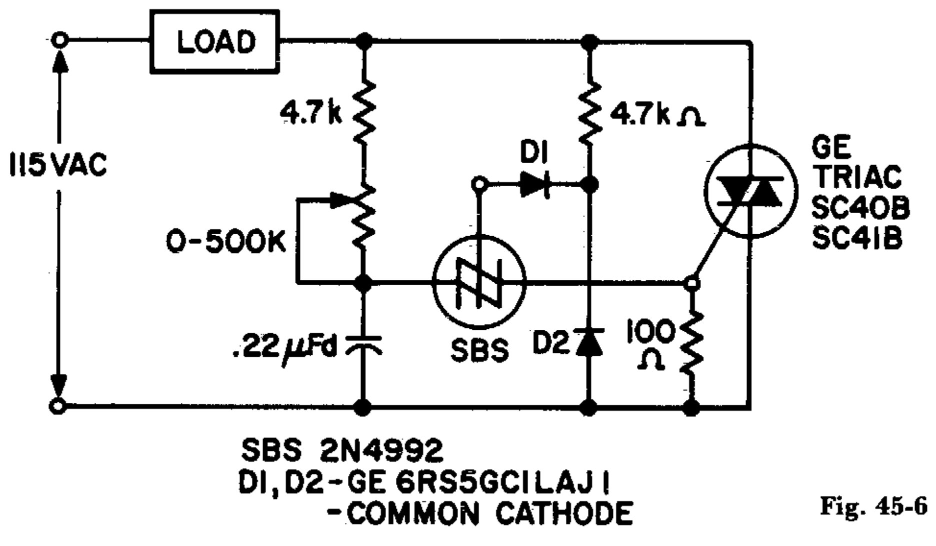


Fig. 45-6

Circuit Notes

This circuit is intended for lamp dimming and similar applications. It requires only one RC phase lag network. To avoid the hysteresis (or "snap-on") effect, the capacitor is reset to approximately 0 volts at the end of every positive half cycle using the gate lead.

LOW COST LAMP DIMMER

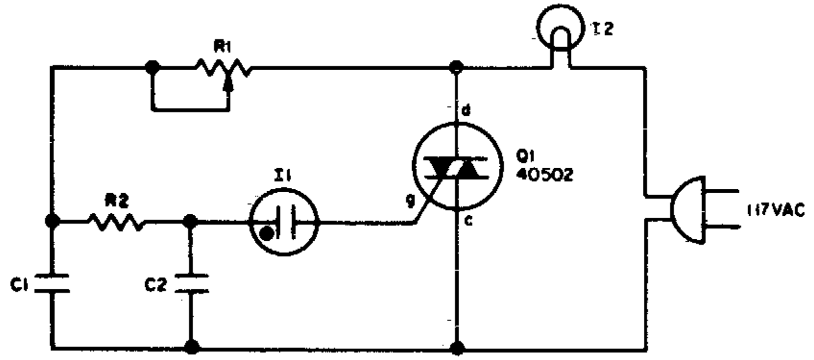


Fig. 45-7

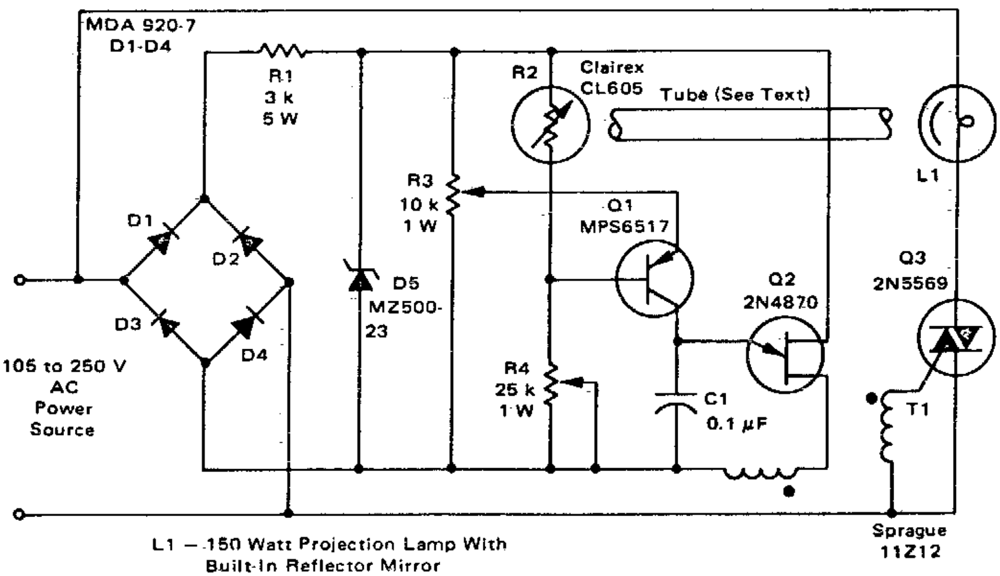
PARTS LIST FOR LO-COST LAMP DIMMER

C1, C2—0.068-uF, 200-VDC capacitor
I1—NE-2 neon lamp
I2—External lamp not to exceed 400 watts
Q1—RCA 40502 Triac
R1—50,000-ohm, pot.
R2—15,000-ohm, ½-watt resistor

Circuit Notes

Without a heatsink, Triac Q1 handles up to a 400-watt lamp. The neon lamp does not trip the gate until it conducts so the lamp turns on a medium brilliance. The lamp can then be backed off to a soft glow.

ZERO-POINT SWITCH



L1 — 150 Watt Projection Lamp With Built-In Reflector Mirror

Fig. 45-8

800 W TRIAC LIGHT DIMMER

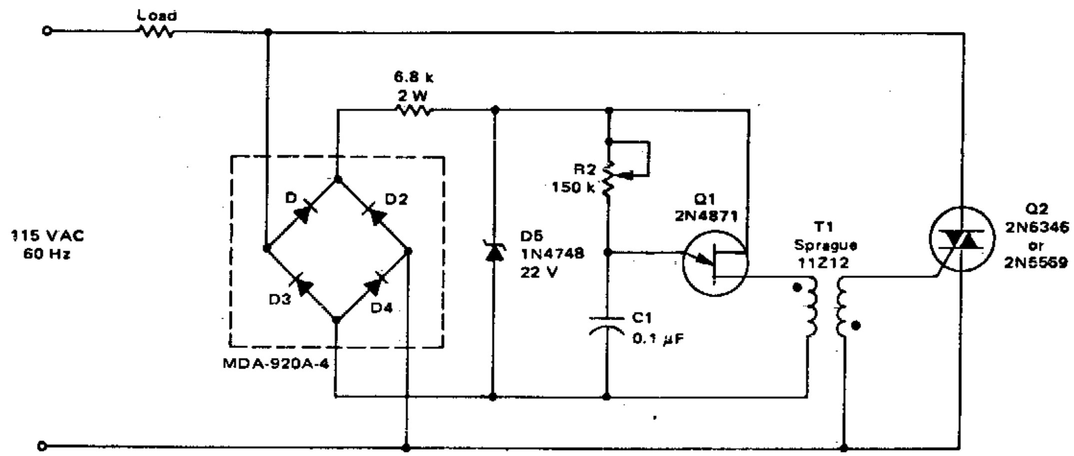
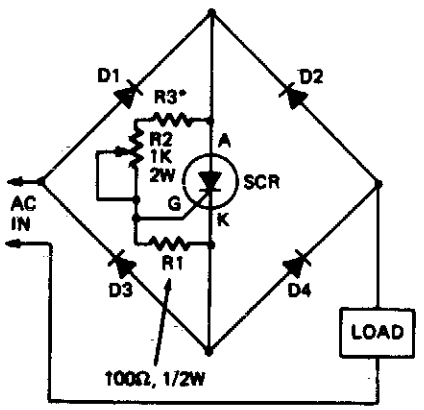


Fig. 45-9

FULL-WAVE SCR CONTROL

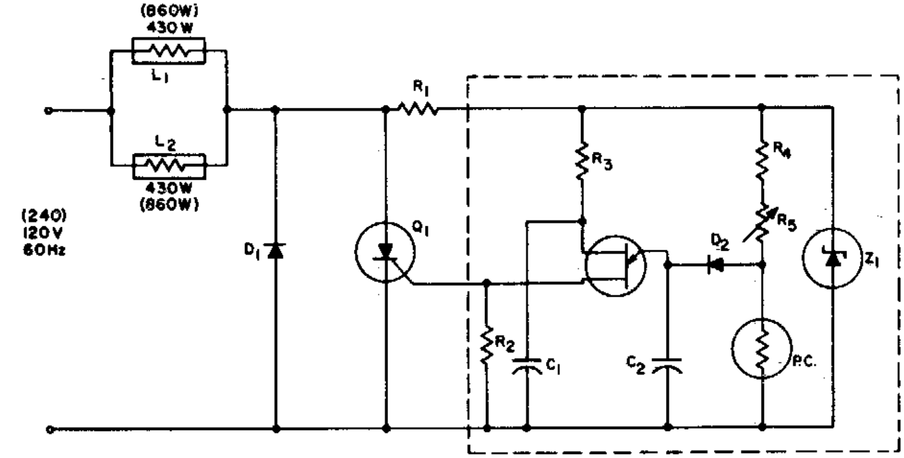


Circuit Notes

This circuit enables a single SCR to provide fullwave control of resistive loads. Resistor R3 should be chosen so that when potentiometer R2 is at its minimum setting, the current in the load is at the required minimum level. Diodes should have same current and voltage rating as the SCR.

Fig. 45-10

860 WATT LIMITED-RANGE LOW COST PRECISION LIGHT CONTROL



R_1 - 6.8KΩ, 2W (15K, 5W)
R_2 - 47Ω
R_3 - 1KΩ
R_4 - 680Ω
R_5 - 1KΩ, 1/2W POT
P.C. - GE A35

Q_1 - GE C122B (C122D)
Q_2 - GE 2N2646
D_1 - GE A41B (A41D)
D_2 - GE A14F
Z_1 - 16 V ZENER
L_1, L_2 - 430W INCANDESCENT LAMP

C_1 - .03μfd, 25V
C_2 - .01μfd, 25V
NOTE: ALL RESISTANCES 1/2w, 10% UNLESS OTHERWISE NOTED.
VALUES IN PARANTHESES FOR 240V

Fig. 45-11

Circuit Notes

The system is designed to regulate an 860 watt lamp load from half to full power. This is achieved by the controlled-half-plus-fixed-half-wave phase control method. Half power applied to an incandescent lamp results in 30% of the full light output. Consequently the circuit is designed to control the light output of the lamp from 30% to 100% of maximum.

800 W SOFT-START LIGHT DIMMER

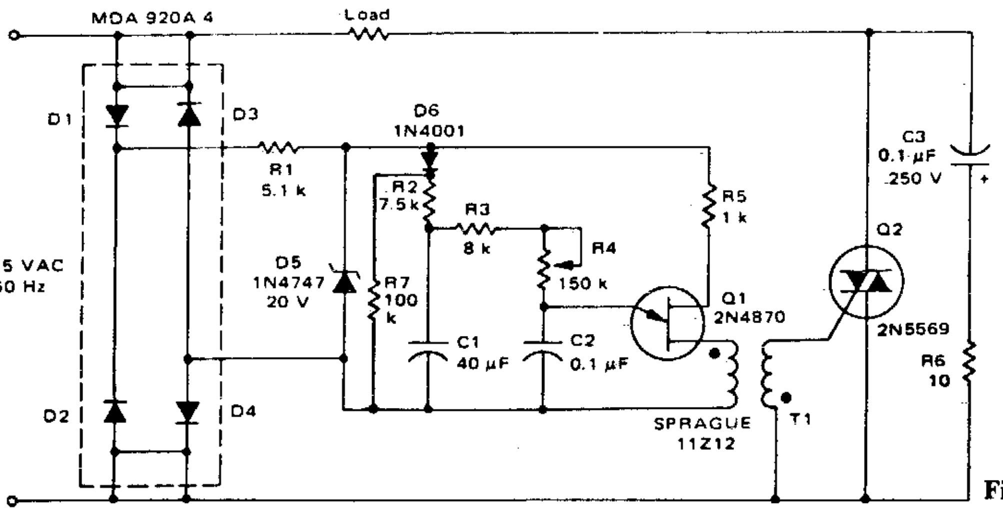


Fig. 45-12

LOW LOSS BRIGHTNESS CONTROL

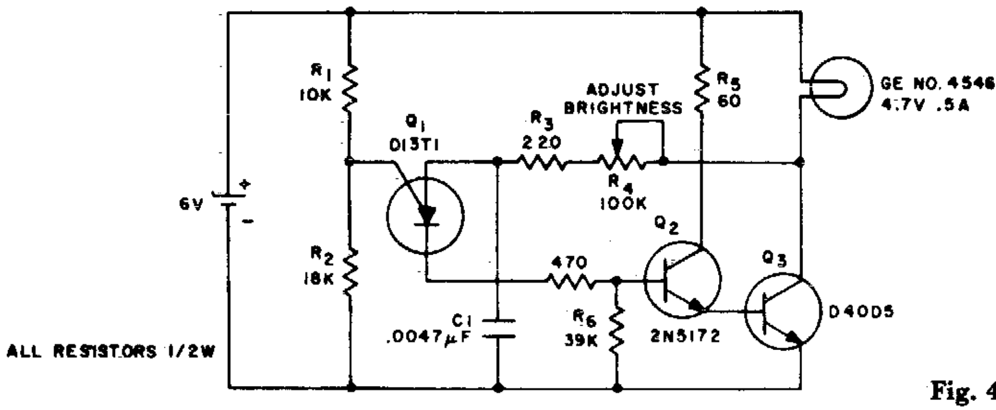


Fig. 45-13

Circuit Notes

This circuit changes the average value of the dc supply voltage because of the high switching frequency. The tungsten lamp will have an almost continuous adjustable light output between 0 and 100%. If a light emitting diode is used as the emitting device, the irradiance will be in phase with the applied current pulses and will decrease to zero when the supply current is zero.

HALF WAVE AC PHASE-CONTROLLED CIRCUIT

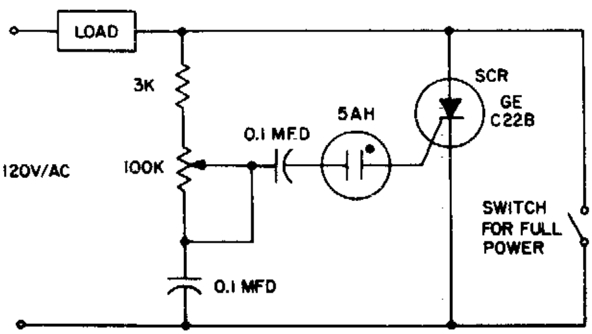


Fig. 45-14

Circuit Notes

The 5AH will trigger when the voltage across the two 0.1 μF capacitors reaches the breakdown voltage of the lamp. Control can be obtained full off to 95% of the half wave RMS output voltage. Full power can be obtained with the addition of the switch across the SCR.

EMERGENCY LIGHT

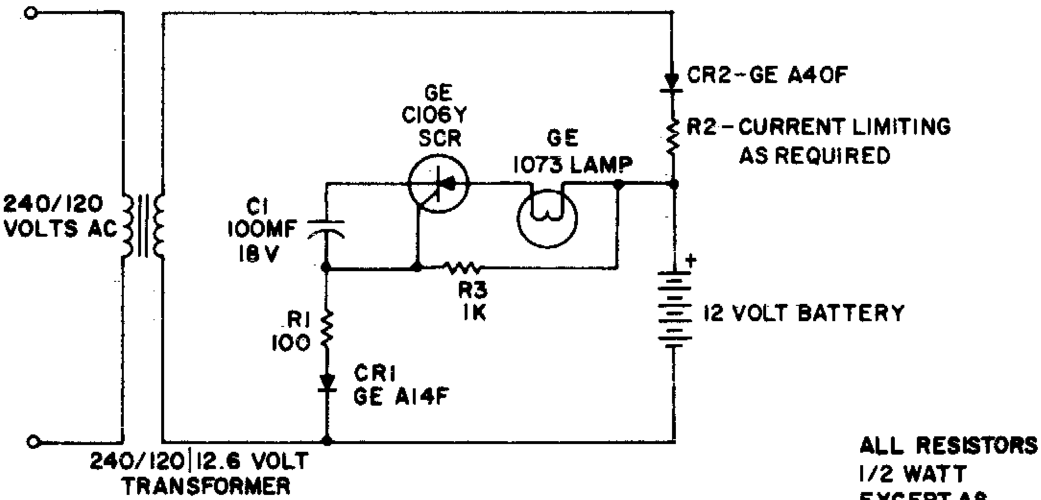


Fig. 45-15

Circuit Notes

This simple circuit provides battery operated emergency lighting instantaneously upon failure of the regular ac service. When line power is restored, the emergency light turns off and the battery recharges automatically. The circuit is ideal for use in elevator cars, corridors and similar places where loss of light due to power failure would be undesirable. Completely static in operation, the circuit requires no maintenance. With ac power on, capacitor C1 charges through rectifier CR1 and resistor R1 to develop a negative voltage at the gate of the C106Y SCR. By this means, the SCR is prevented from being triggered, and the emergency light stays off. At the same time, the battery is kept fully charged by rectifier CR2 and resistor R2. Should the ac power fail, C1 discharges and the SCR is triggered on by battery power through resistor R3. The SCR then energizes the emergency light. Reset is automatic when ac is restored, because the peak ac line voltage biases the SCR and turns it off.

NEON LAMP DRIVER

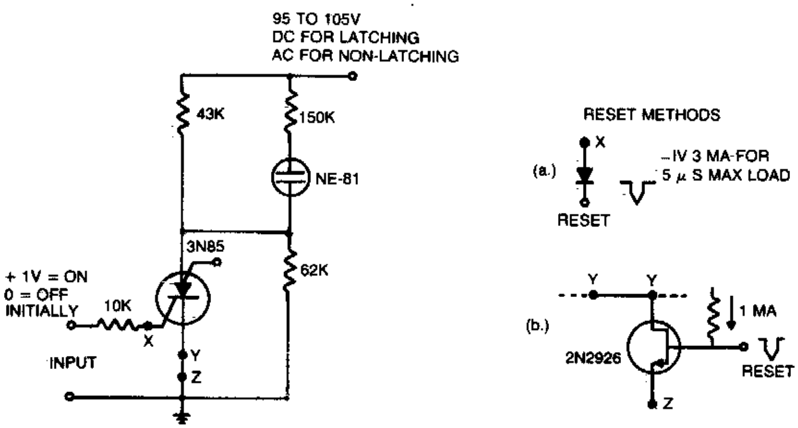


Fig. 45-16

COMPLEMENTARY AC POWER SWITCHING

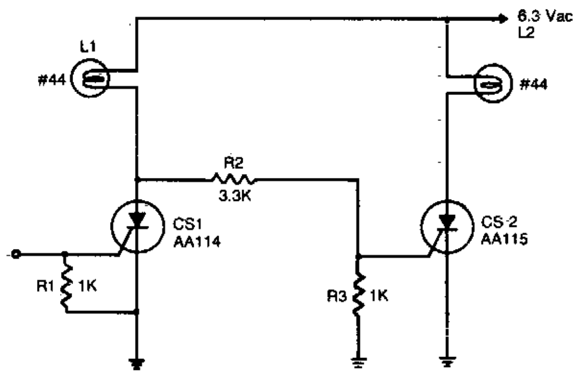


Fig. 45-17

Circuit Notes

An input signal of less than 1 mA and 1 V is required to switch on CS1. As long as this input signal is maintained, CS1 will conduct during each positive half cycle of anode voltage, thereby energizing load L1 with half-wave rectified dc. L2 remains de-energized, since the anode of CS1 will not go more positive than 1.5 volts, and voltage divider R2 - R3 cannot provide enough voltage to trigger CS2. Upon removal of the input signal, CS1 will drop out. L1 will be de-energized, except for a small amount of ac current through R2 and R3. CS2 will be triggered on at the beginning of each positive half-cycle, when CS1 anode voltage reaches 2 to 3 volts. CS2 will conduct for nearly the entire positive half-cycle energizing L2. It should be noted that the 6.3 volt lamps used will operate at ⅓ the rated brilliance because of the controlled switch half-wave rectifying action and will extend the operating lamp life by several orders of magnitude. Should full brilliance be desired, the anode supply voltage level should be raised to 9 volts ac.

BATTERY LANTERN CIRCUIT

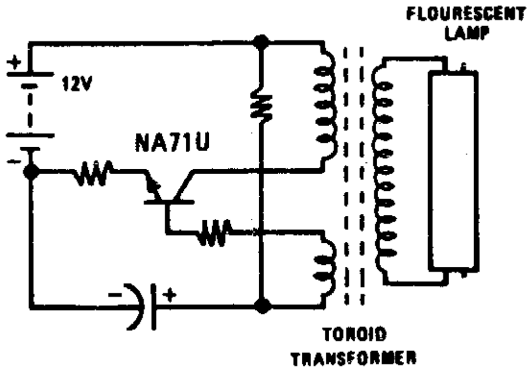
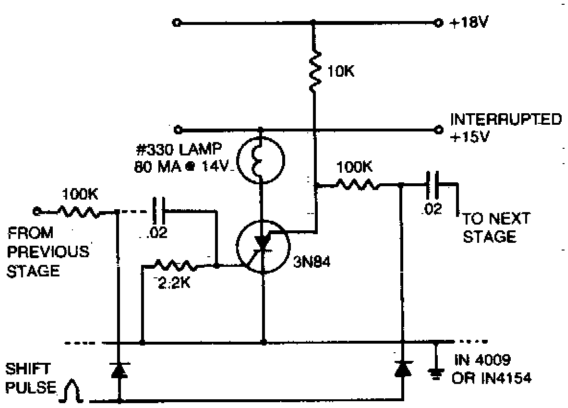


Fig. 45-18

SHIFT REGISTER



BASIC SHIFT REGISTER STAGE

Fig. 45-19

Circuit Notes

The shift pulse amplitude is less than 15 volts. If a stage is off, the shift pulse will not be coupled to the next stage. If it is on, the diode will conduct and trigger the next stage. Just prior to the shift pulse the anode supply is interrupted to turn off all stages. The stored capacitor charge determines which stages will be triggered.

LIGHT-LEVEL CONTROLLER

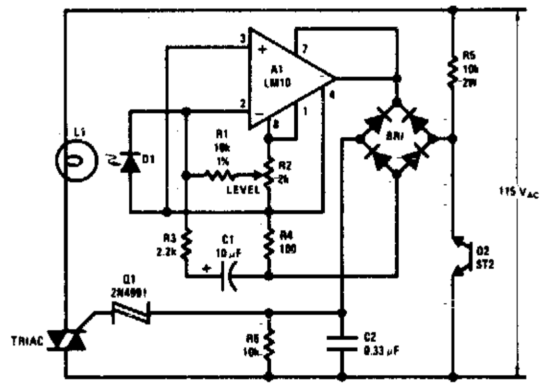


Fig. 45-20

2.2 WATT INCANDESCENT LAMP DRIVER

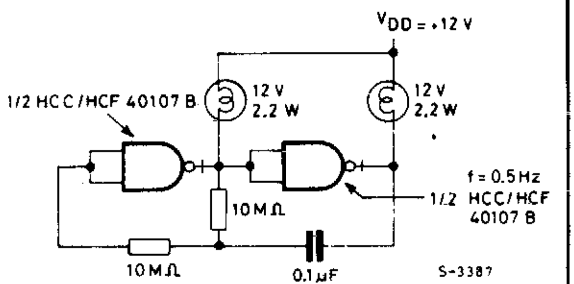


Fig. 45-21

46

Light Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LINEAR LIGHT-METER CIRCUIT

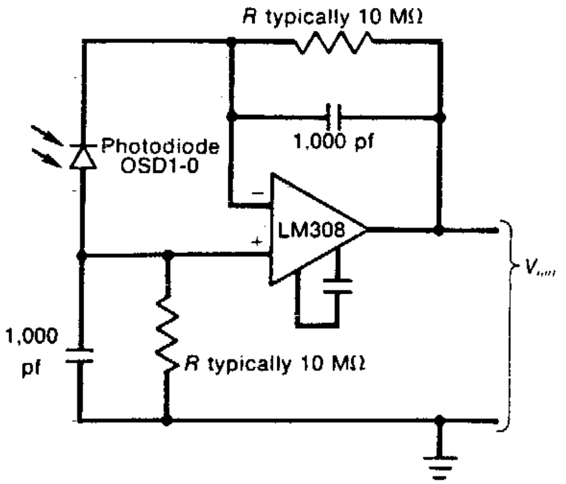


Fig. 46-1

Circuit Notes

This circuit uses a low-input-bias op amp to give a steady dc indication of light level. To reduce circuit sensitivity to light, R1 can be reduced, but should not be less than 100 K. The capacitor values in the circuit are chosen to provide a time constant sufficient to filter high-frequency light variations that might arise, for example, from fluorescent lights.

LOGARITHMIC LIGHT-METER CIRCUIT

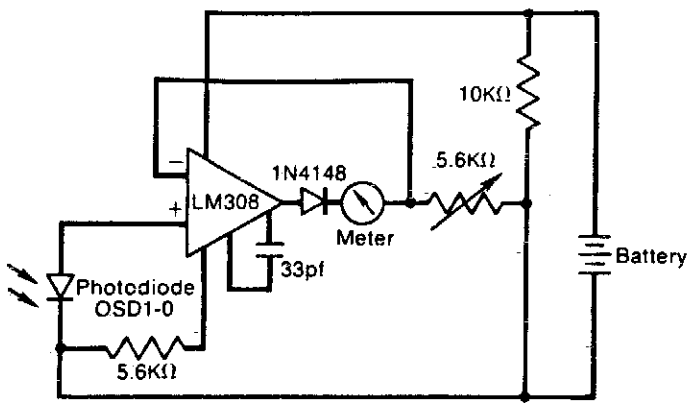


Fig. 46-2

Circuit Notes

The meter reading is directly proportional to the logarithm of the input light power. The logarithmic circuit behavior arises from the nonlinear diode pnjunction current/voltage relationship. The diode in the amplifier output prevents output voltage from becoming negative (thereby pegging the meter), which may happen at low lightlevels due to amplifier bias currents. R1 adjusts the meter full-scale deflection, enabling the meter to be calibrated.

LIGHT METER

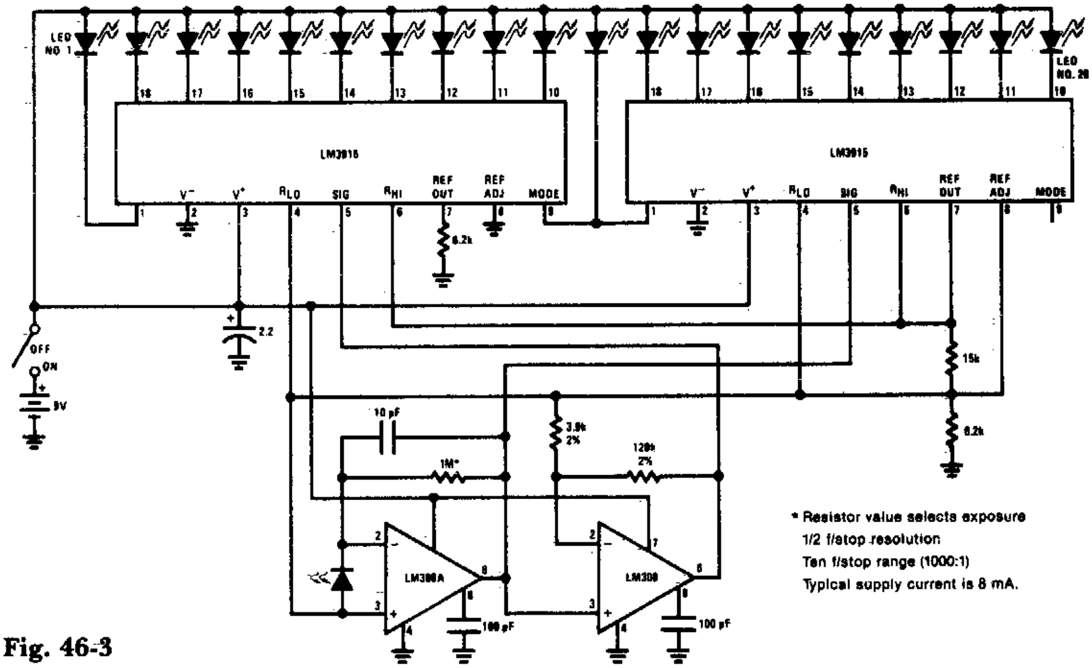


Fig. 46-3

LIGHT METER

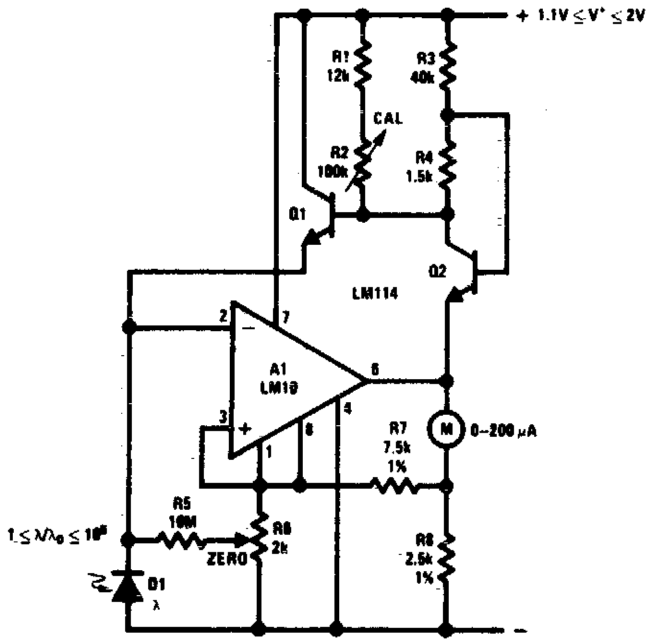


Fig. 46-4

LIGHT METER

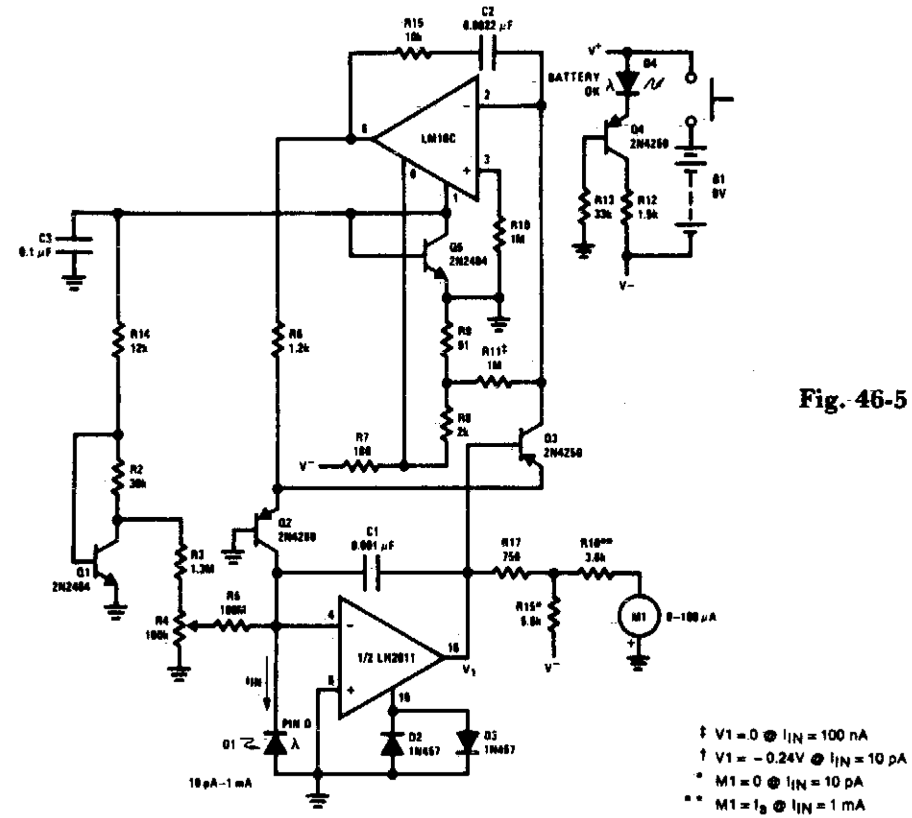


Fig. 46-5

Circuit Notes

This light meter has an eight-decade range. Bias current compensation can give input current resolution of better than ±2 pA over 15 °C to 55 °C.

PRECISION PHOTODIODE COMPARATOR

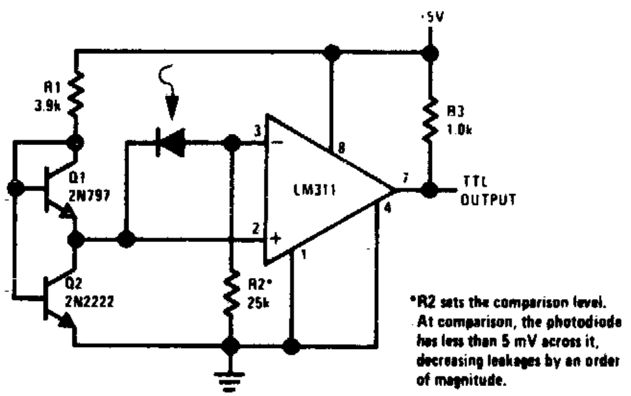


Fig. 46-6

47

Liquid Level Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LEVEL SENSOR FOR CRYOGENIC FLUIDS

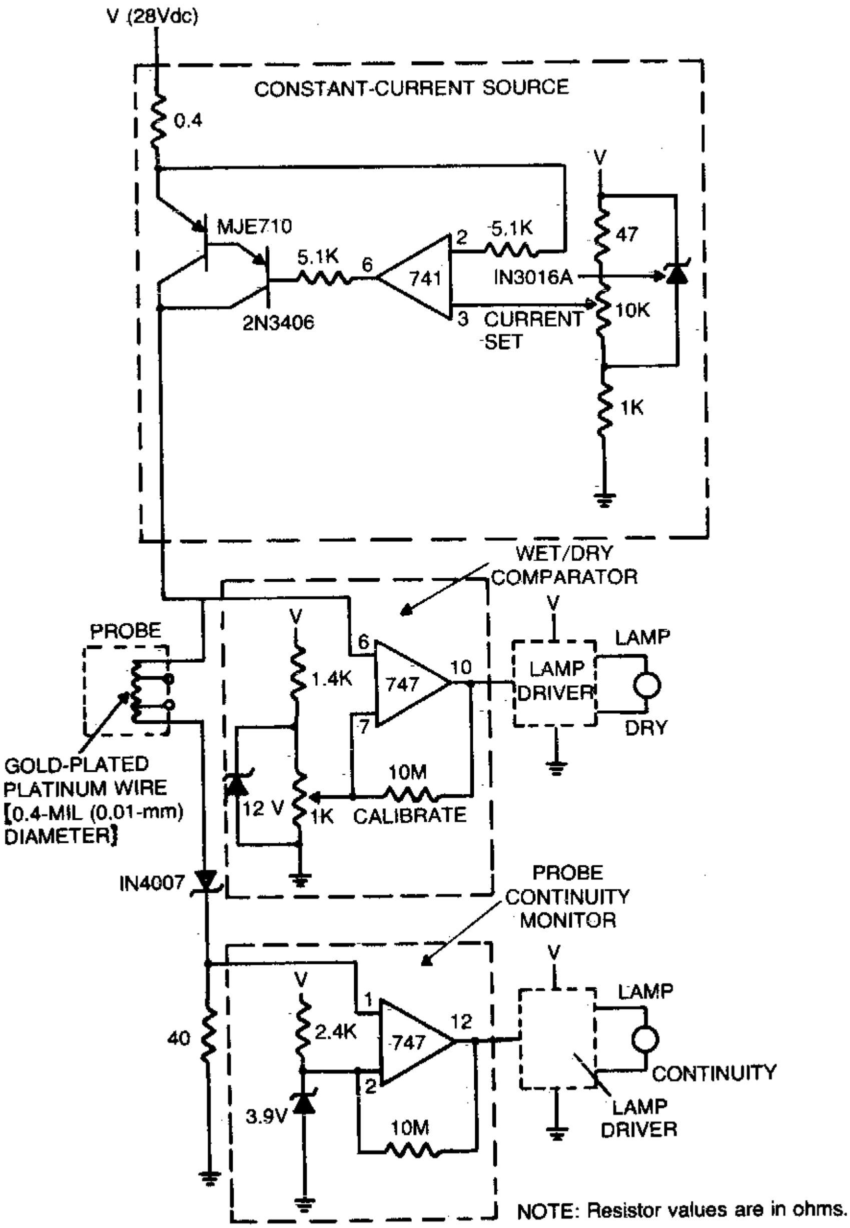


Fig. 47-1

Circuit Notes

The sensor circuit is adaptable to different liquids and sensors. The constant-current source drives current through the sensing probe and a fixed resistor. The voltage-comparator circuits interpret the voltage drops to tell whether the probe is immersed in liquid and whether there is current in the probe.

FLUID LEVEL CONTROLLER

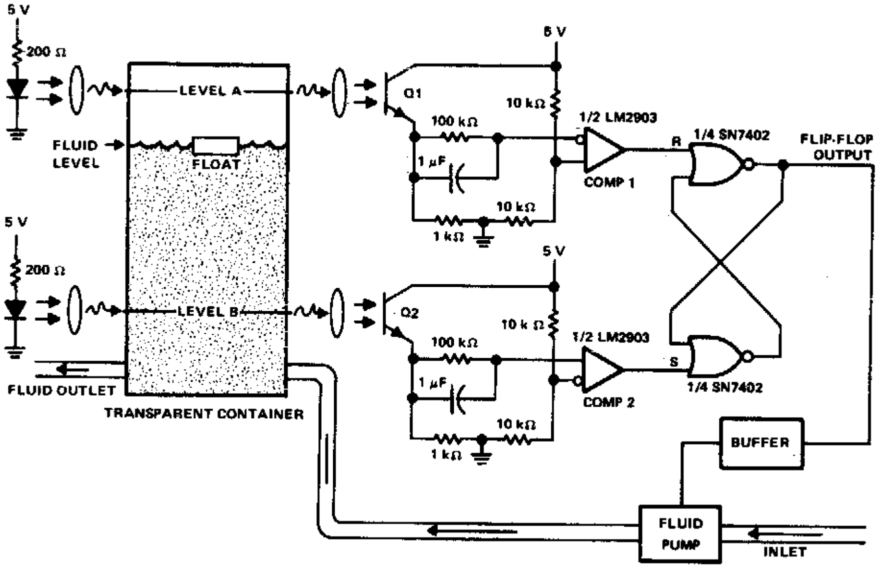
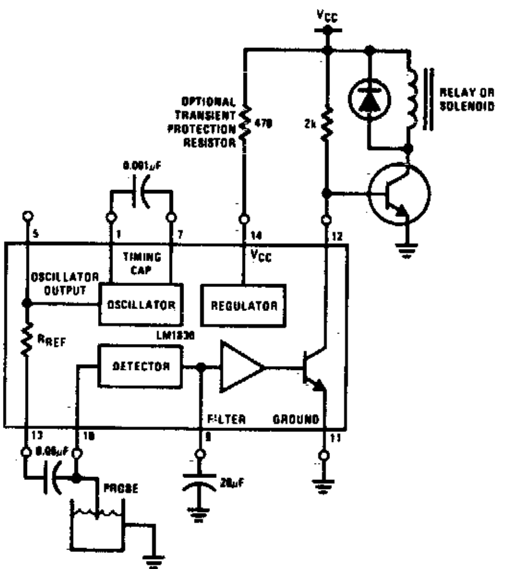


Fig. 47-2

Circuit Notes

This circuit can be used to maintain fluid between two levels. Variations on this control circuit can be made to keep something that moves within certain boundary conditions.

HIGH LEVEL WARNING DEVICE



The output is suitable for driving a sump pump or opening a drain valve, etc.

Fig. 47-3

LIQUID LEVEL CONTROL

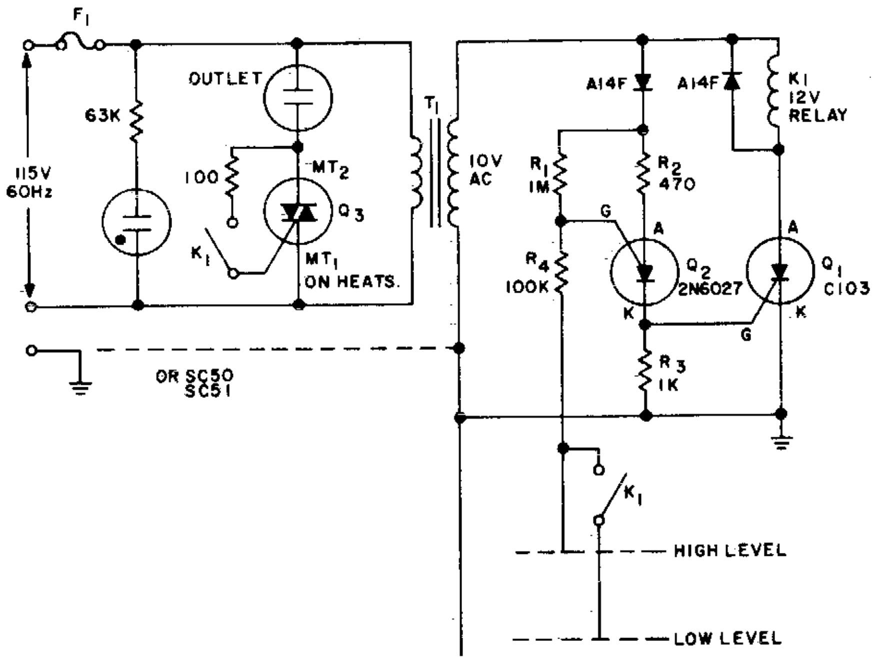
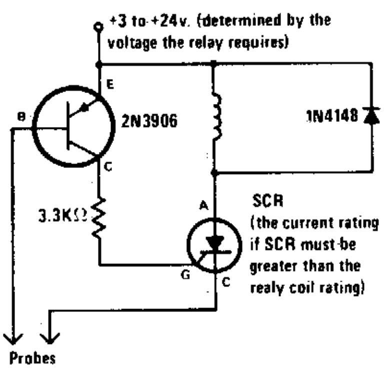


Fig. 47-4

Circuit Notes

Use this circuit to keep the fluid level of a liquid between two fixed points. Two modes, for filling or emptying are possible by simple reversing the contact connections of K1. The loads can be either electric motors or solenoid operated valves, operating from ac power. Liquid level detection is accomplished by two metal probes, one measuring the high level and the other the low level. An inversion of the logic (keeping the container filled) can be accomplished by replacing the normally open contact on the gate of Q3 with a normally closed contact.

LIQUID LEVEL DETECTOR (LATCHING)



Circuit Notes

Alarm is actuated when liquid level is above the probes and remains activated even if the level drops below the probes. This latching action lets you know that the pre-set level has been reached or exceeded sometime in the past.

Fig. 47-5

WATER LEVEL ALARM

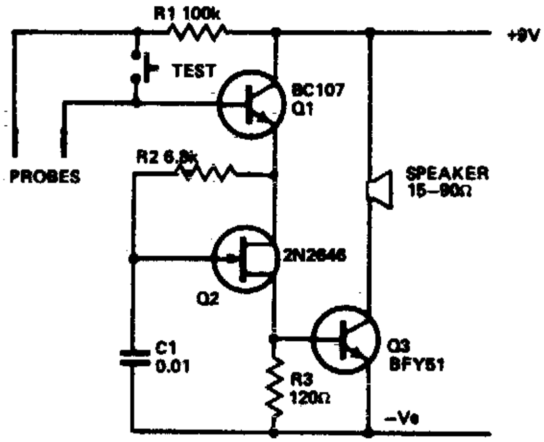


Fig. 47-6

Circuit Notes

The circuit draws so little current that the shelf-line of the battery is the limiting factor. The only current drawn is the leakage of the transistor. The circuit is shown in the form of a water level alarm but by using different forms of probe can act as a rain alarm or shorting alarm; anything from zero to about 1 M between the probes will trigger it. Q1 acts as a switch which applies current to the unijunction relaxation oscillator Q2. Alarm signal frequency is controlled by values and ratios of C1/R2. Pulses switch Q3 on and off, applying a signal to the speaker. Almost any NPN silicon transistor can be used for Q1 and Q3 and almost any unijunction for Q2.

WATER-LEVEL SENSING CONTROL CIRCUIT

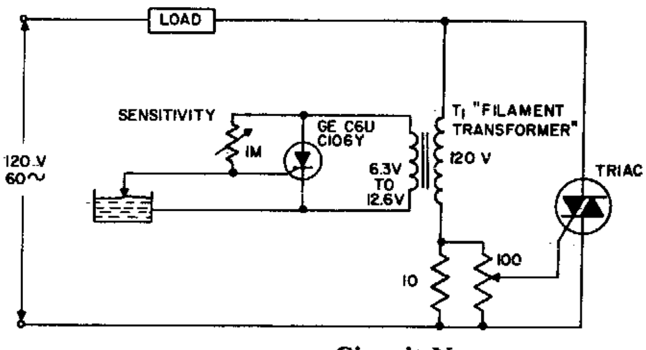
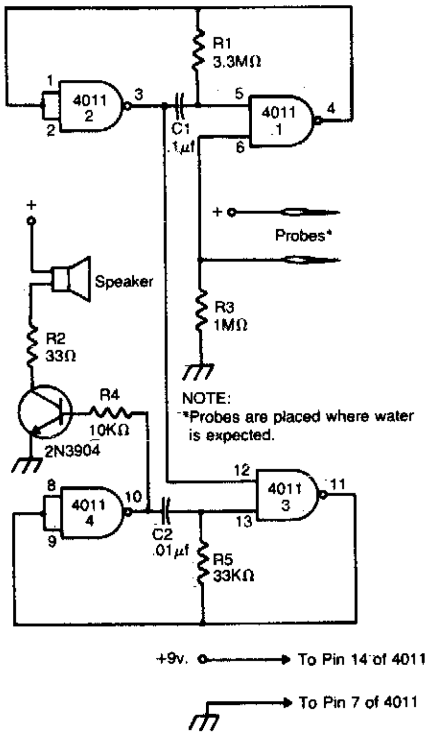


Fig. 47-7

Circuit Notes

The circuit applies power to the load until the water conducts through the probe, and bypasses gate current from the low current SCR. This gives an isolated low voltage probe to satisfy safety requirements.

FLOOD ALARM

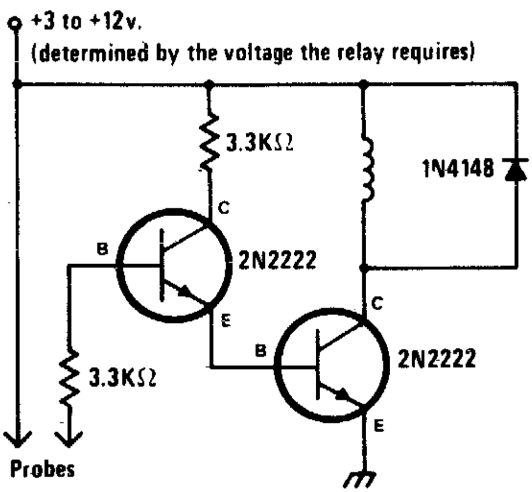


Circuit Notes

The alarm is built around two audio oscillators, each using two NAND gates. The detection oscillator is gated on by a pair of remote probes. One of the probes is connected to the battery supply, the other to the input of one of the gates. When water flows between the probes, the detection oscillator is gated on. The alarm oscillator is gated on by the output of the detection oscillator. The values given produce an audio tone of about 3000 Hz. The detection oscillator gates this audio tone at a rate of about 3 Hz. The result is a unique pulsating note. Use any 8 ohm speaker to sound the alarm. The 2N3904 can be replaced by any similar NPN transistor. The circuit will work from any six to 12-volt supply.

Fig. 47-8

LIQUID LEVEL DETECTOR



Circuit Notes

When liquid level reaches both probes, alarm is turned on. When water level recedes it goes off.

Fig. 47-9

LOW-LEVEL WARNING WITH AUDIO OUTPUT

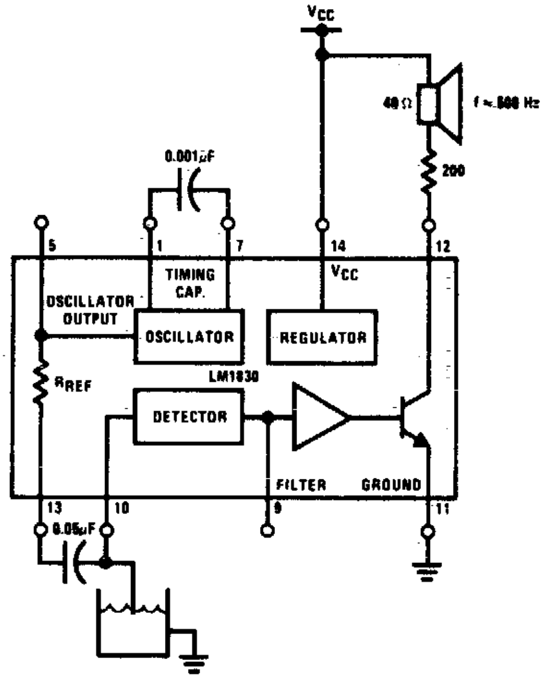


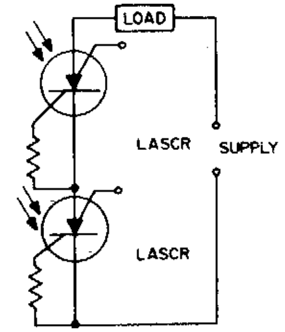
Fig. 47-10

48

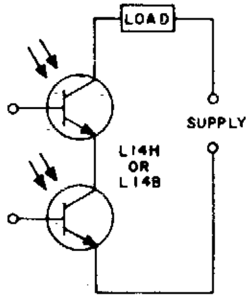
Logic Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

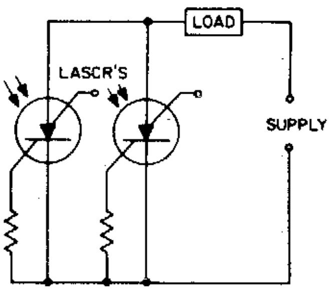
LIGHT ACTIVATED LOGIC CIRCUITS



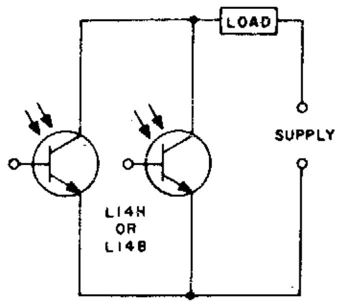
(a) AND Circuit



(b) AND Circuit

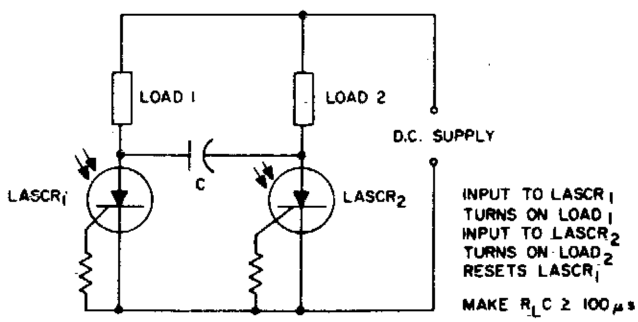


(c) OR Circuit



(d) OR Circuit

Fig. 48-1

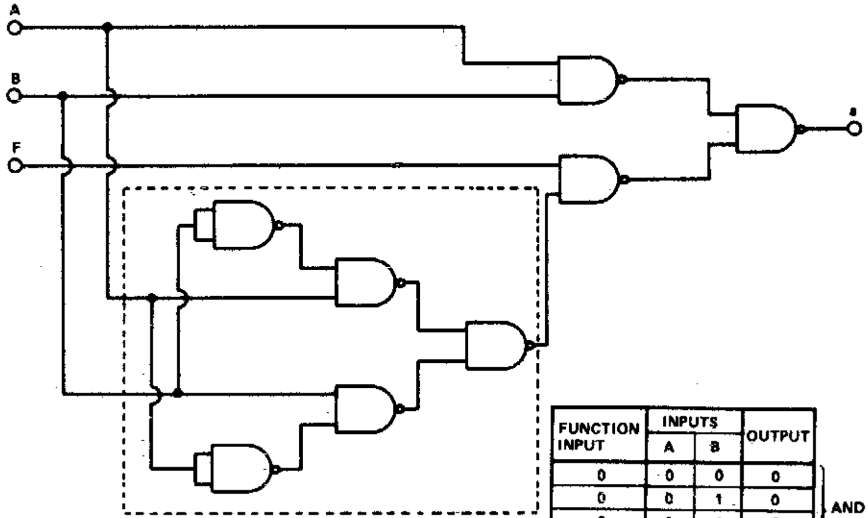


(e) Flip-Flop

Circuit Notes

These circuits illustrate some of the common logic functions that can be implemented.

PROGRAMMABLE GATE



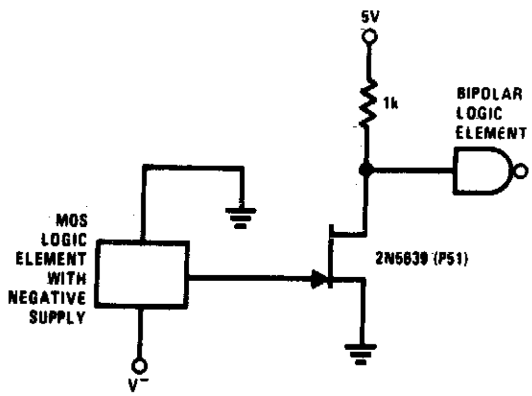
FUNCTION INPUT	INPUTS A	INPUTS B	OUTPUT	
0	0	0	0	AND FUNCTION
0	0	1	0	
0	1	0	0	
0	1	1	1	
1	0	0	0	OR FUNCTION
1	0	1	1	
1	1	0	1	
1	1	1	1	

Fig. 48-2

Circuit Notes

This gate converts an AND gate or an OR gate by applying a logic '1' on the function input. The logic design uses 8 two-input NAND gates. The number of gates may be reduced by replacing the 5 NAND gates enclosed by the dotted line with a two-input exclusive-OR, such as the TTL 7486.

NEGATIVE TO POSITIVE SUPPLY LOGIC LEVEL SHIFTER



Circuit Notes

This simple circuit provides for level shifting from any logic function (such as MOS) operating from minus to ground supply to any logic level (such as TTL) operating from a plus to ground supply. The 2N5639 provides a low $r_{dc\,(ON)}$ and fast switching times.

Fig. 48-3

OR GATE

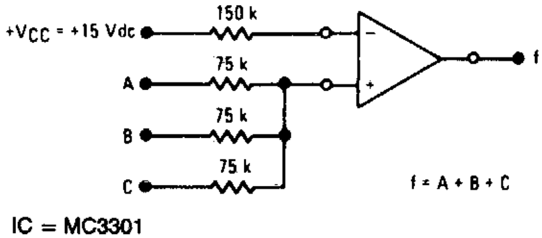


Fig. 48-4

AND GATE

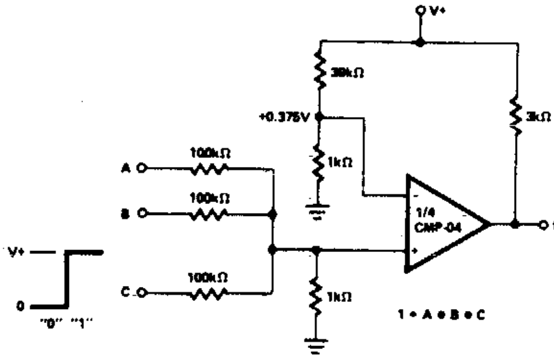


Fig. 48-7

OR GATE

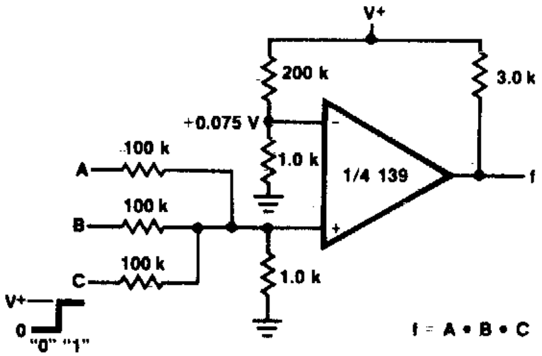


Fig. 48-5

R-S FLIP-FLOP

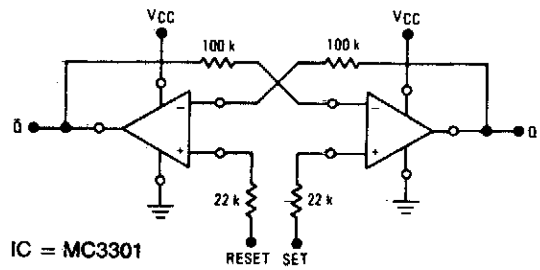


Fig. 48-8

LARGE FAN-IN AND GATE

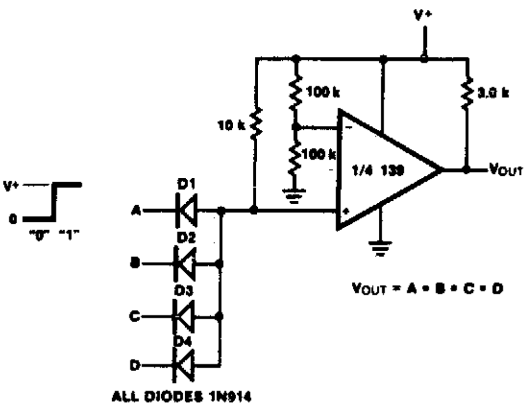


Fig. 48-6

AND GATE

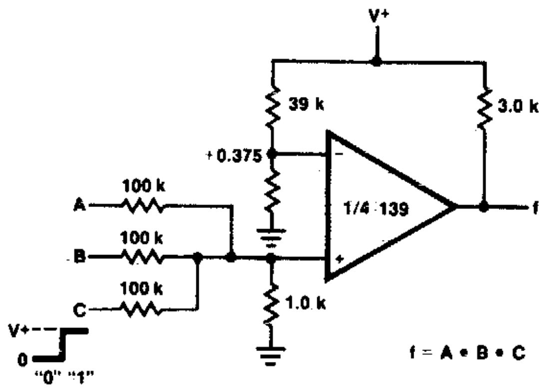


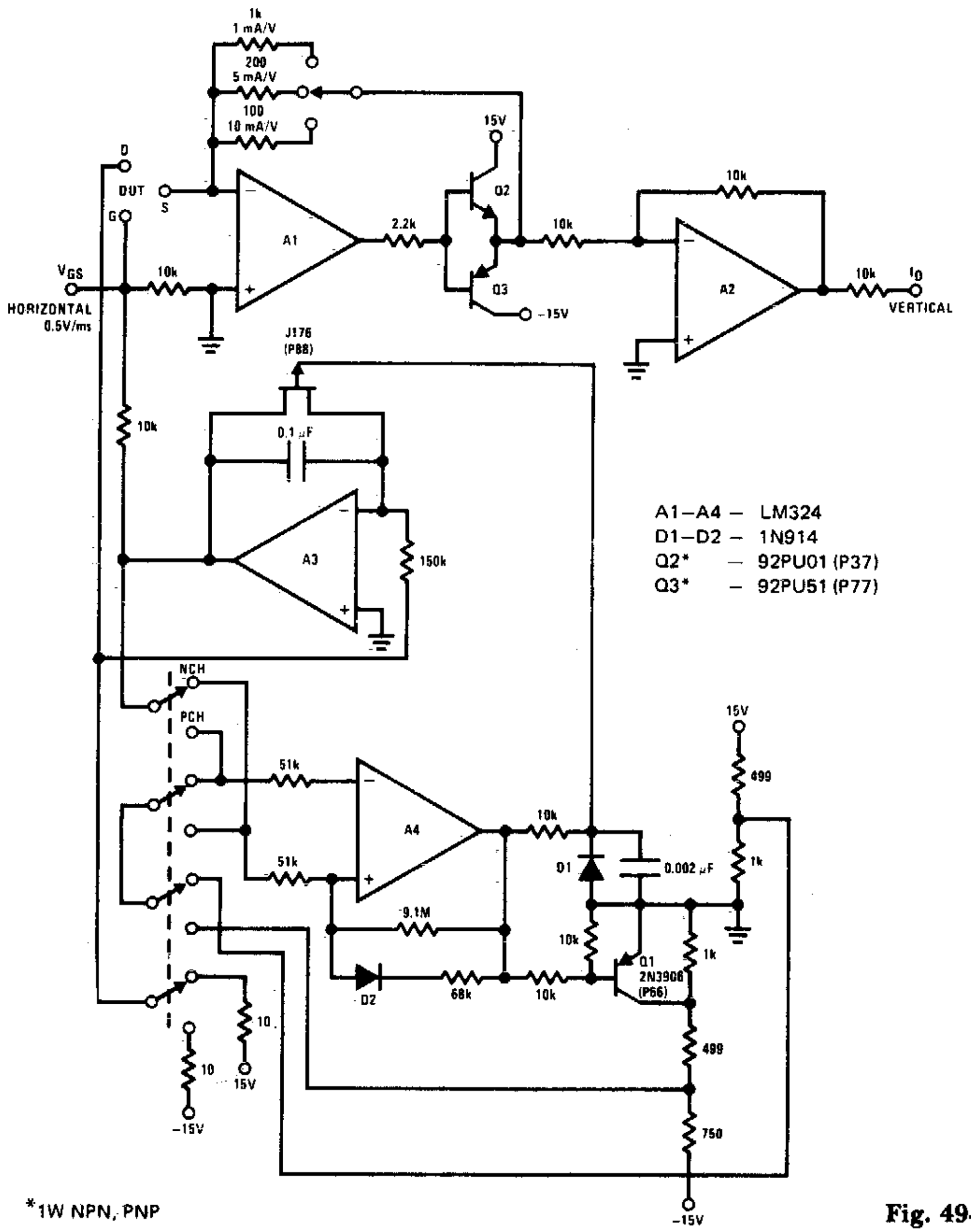
Fig. 48-9

49

Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FET CURVE TRACER



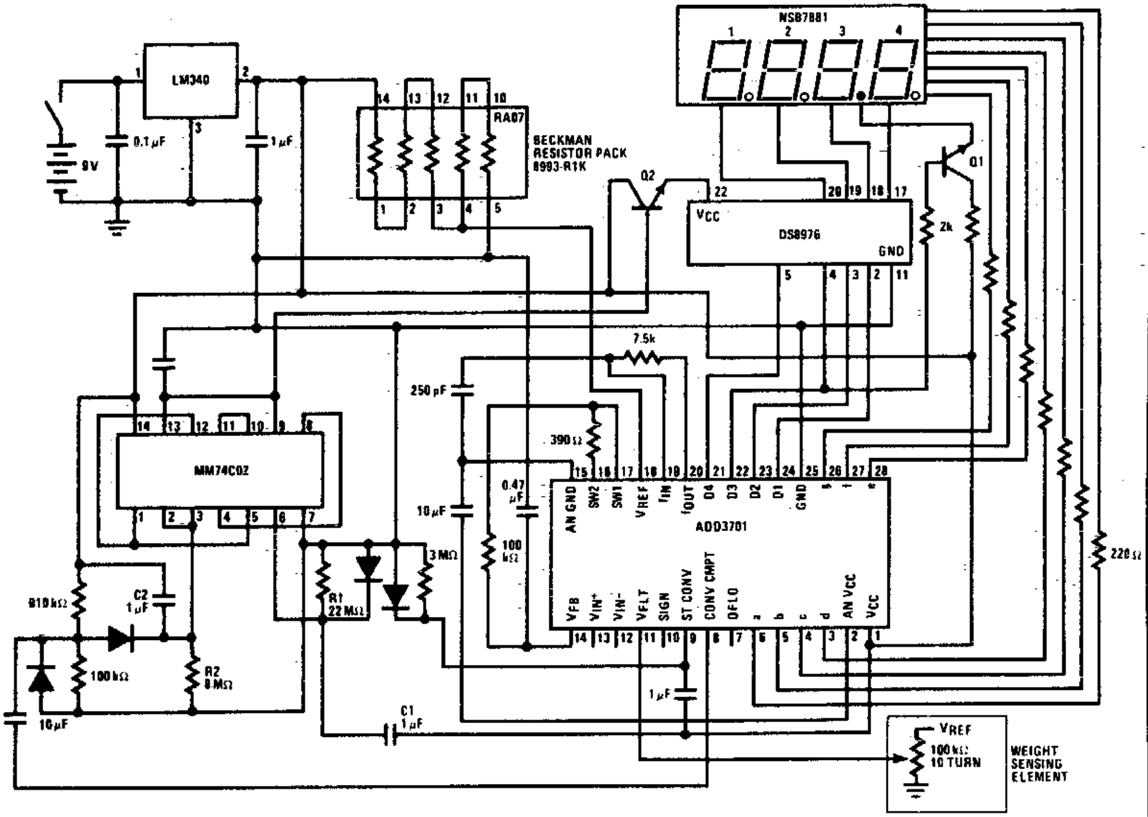
*1W NPN, PNP

Fig. 49-1

Circuit Notes

The circuit displays drain current versus gate voltage for both P and N-channel JFETS at a constant drain voltge.

DIGITAL WEIGHT SCALE



Notes:

1. R1, C1 defines POWER ON display blanking interval. R2, C2 defines display ON time.
2. All V_{CC} connections should use a single V_{CC} point and all ground/analog ground connections should use a single ground/analog ground point.
3. Display sequence for Rev A ckt implementation:

t = 0 sec	• power ON
t = 0 → 5 sec	• display blanked
	• system converging
t = 5 → 10 sec	• conversion complete
	• display ENABLE
t ⩾ 10 sec	• display blanked
	• wait for new POWER UP cycle

Fig. 49-2

Circuit Notes

This circuit employs a potentiometer as the weight sensing element. An object placed upon the scale displaces the potentiometer wiper, an amount proportional to its weight. Conversion of the wiper voltage to digital information is performed, decoded, and interfaced to the numeric display.

LOW COST pH METER

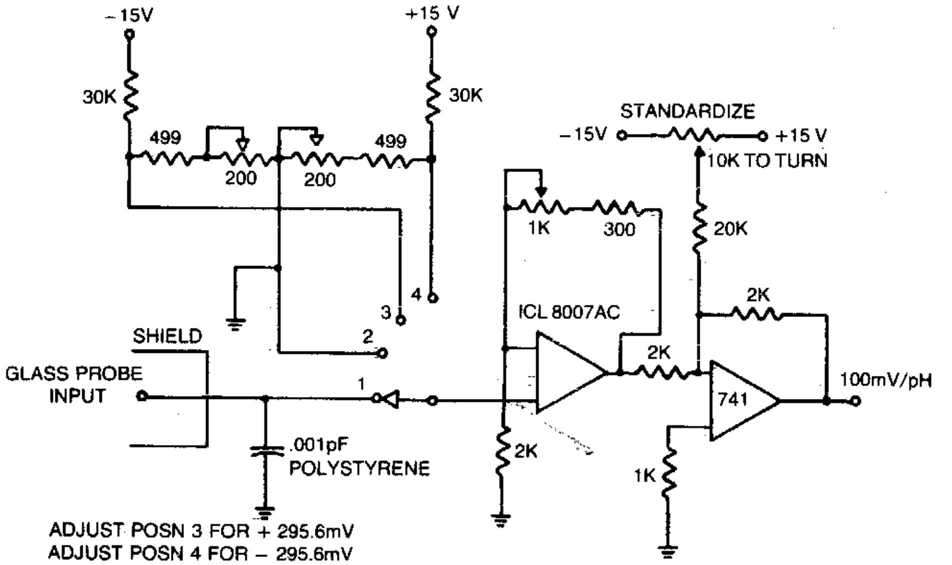
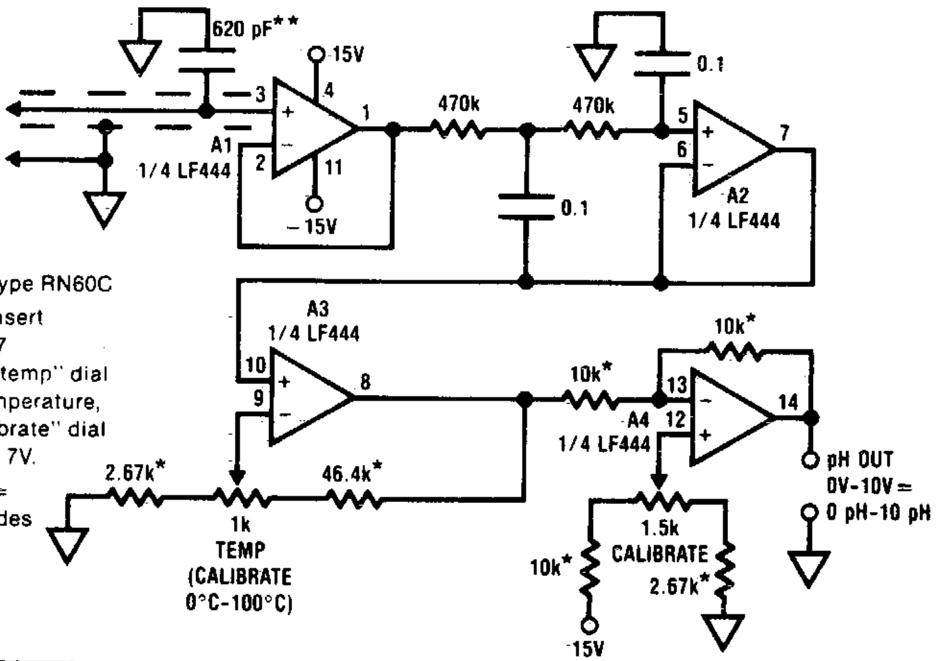


Fig. 49-3

Circuit Notes

With guaranteed 1 pA input bias, the ICL 8007A is ideal as a pH meter or long term sample and hold.

pH PROBE AMPLIFIER/TEMPERATURE COMPENSATOR



** Polystyrene

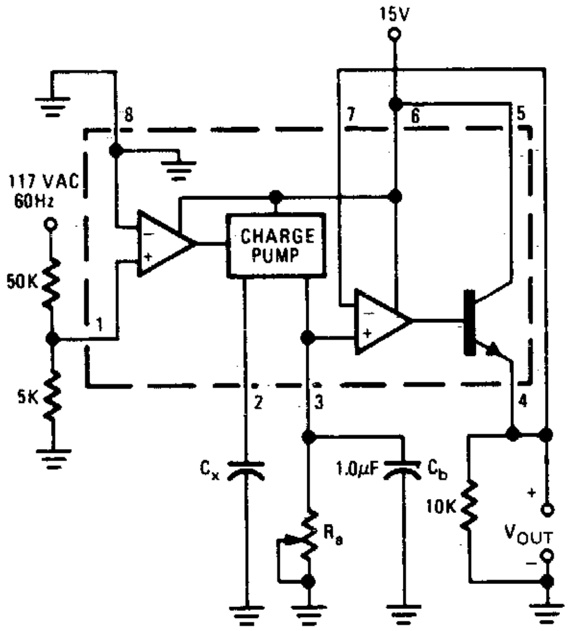
* Film resistor type RN60C

To calibrate, insert probe in pH = 7 solution. Set "temp" dial to solution temperature, then, set "calibrate" dial so output read 7V.

Typical probe = Ingold Electrodes #465-35

Fig. 49-4

CAPACITANCE METER



Circuit Notes

Output voltage is proportional to the capacitance connected to pin 2 of the charge pump. The meter works over a range of 0.01 to 0.1 μF with R_a set at 111 K. Over this range of capacitance, the output voltage varies from 1 to 10 volts with a 15 volt power supply. A constant frequency reference is taken from the 60-Hz line.

Fig. 49-5

ZENER TESTER

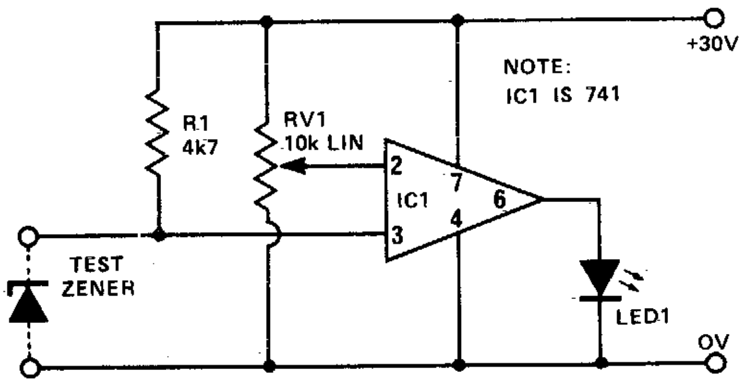


Fig. 49-6

Circuit Notes

This circuit provides a low cost and reliable method of testing zener diodes. RV1 can be calibrated in volts, so that when LED 1 just lights, the voltage on pins 2 and 3 are nearly equal. Hence, the zener voltage can be read directly from the setting of RV1. The supply need only be as high a value as the zener itself. For a more accurate measurement, a precision pot could be added and calibrated.

TRANSISTOR SORTER/TESTER

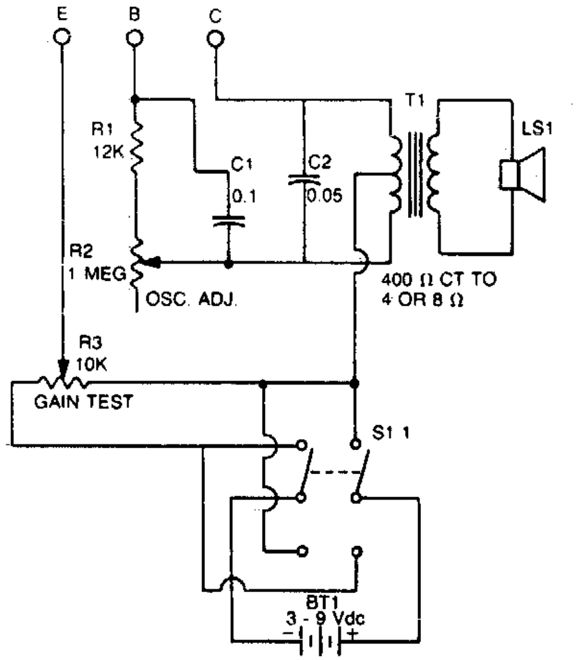


Fig. 49-7

Circuit Notes

This tester checks transistor for polarity (PNP or NPN). An audible signal will give an indication of gain. Tester can also be used as a GO/NO GO tester to match unmarked devices.

GO/NO-GO DIODE TESTER

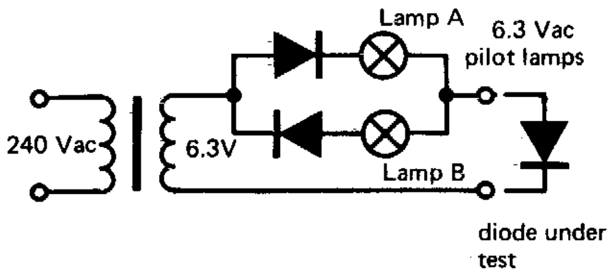
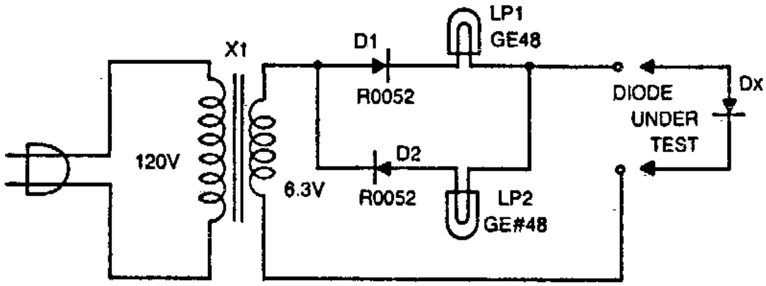


Fig. 49-8

Circuit Notes

If lamp A or B is illuminated, the diode is serviceable. If both light, the diode is short circuited. If neither light, diode is an open circuit.

DIODE TESTER



Parts list
D1, D2—HEP R0052
LP1, LP2—GE 48 lamp
X1—120V to 6.3V, transformer, Lafayette 33P80508

Fig. 49-9

Circuit Notes

The circuit tests whether or not a diode is open, shorted, or functioning correctly. If lamp A lights, the diode under test is functional. When lamp B is lit, the diode is good but connected backwards. When both lamps are lit, the diode is shorted, and it is open if neither lamp is lit.

PEAK LEVEL INDICATOR

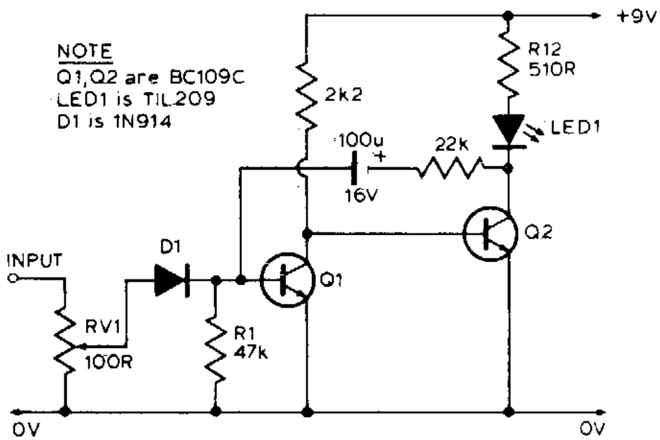


Fig. 49-10

Circuit Notes

The LED is normally lit, but it will be briefly extinguished if the input exceeds a preset (by RV1) level. A possible application is to monitor the output voltage across a loudspeaker; the LED will flicker with large signals.

SOUND LEVEL MONITOR

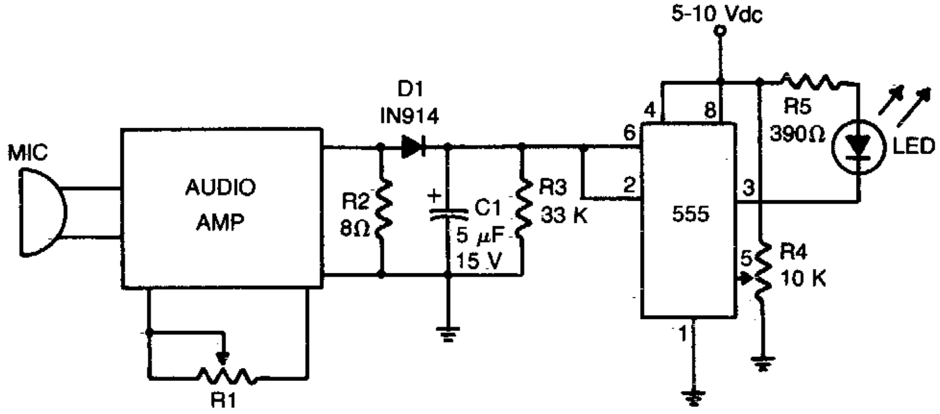


Fig. 49-11

Circuit Notes

Loudness detector consists of a 555 IC wired as a Schmitt trigger. The output changes state—from high to low—whenever the input crosses a certain voltage. That threshold voltage is established by the setting of R4.

LINEAR VARIABLE DIFFFERENTIAL TRANSFORMER (LVDT) DRIVER DEMODULATOR

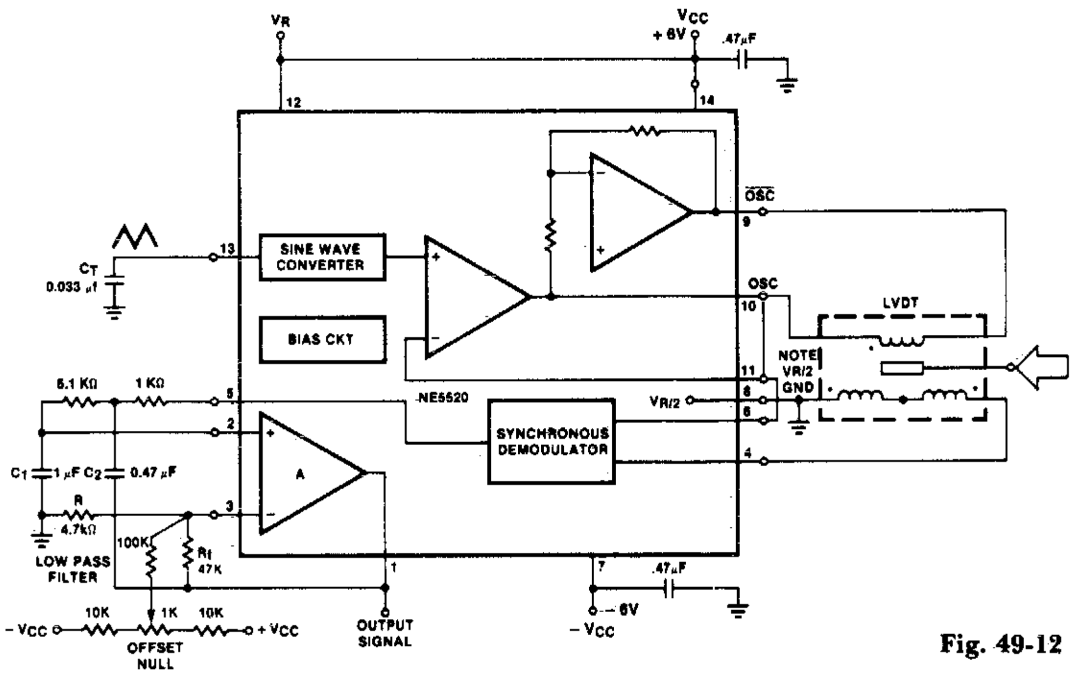


Fig. 49-12

LINEAR VARIABLE DIFFERENTIAL TRANSFORMER (LVDT) MEASURING GAUGE

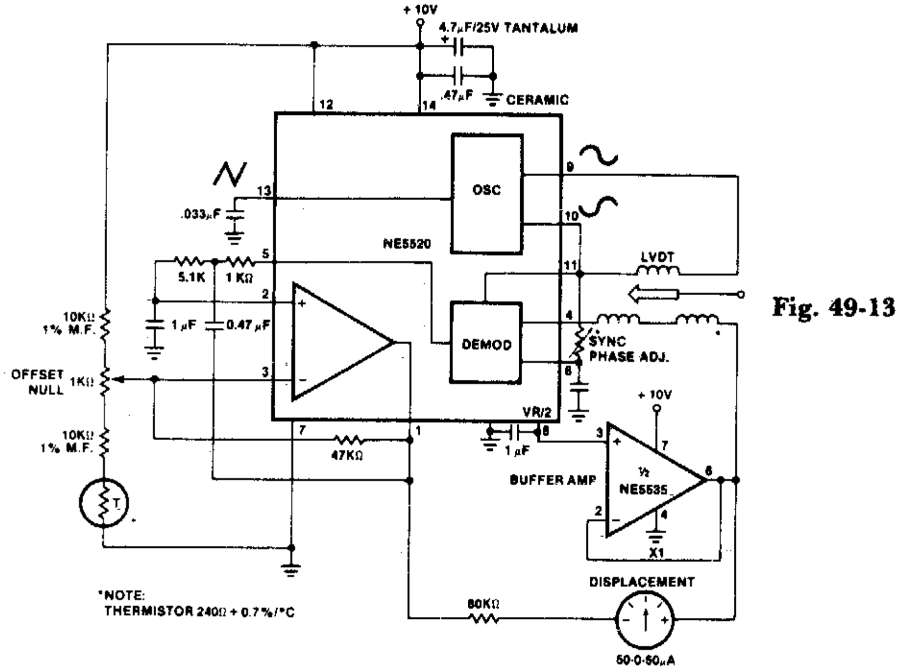
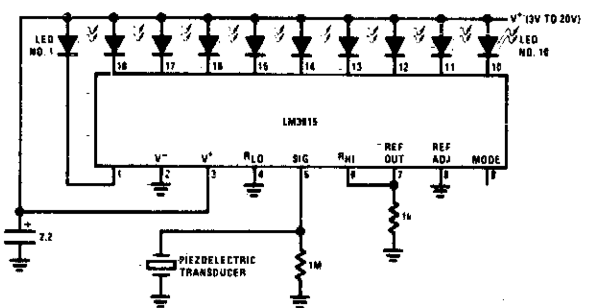


Fig. 49-13

VIBRATION METER



LED	Threshold
1	60 mV
2	80 mV
3	110 mV
4	160 mV
5	220 mV
6	320 mV
7	440 mV
8	630 mV
9	890 mV
10	1.25 V

Fig. 49-14

SENSITIVE RF VOLTMETER

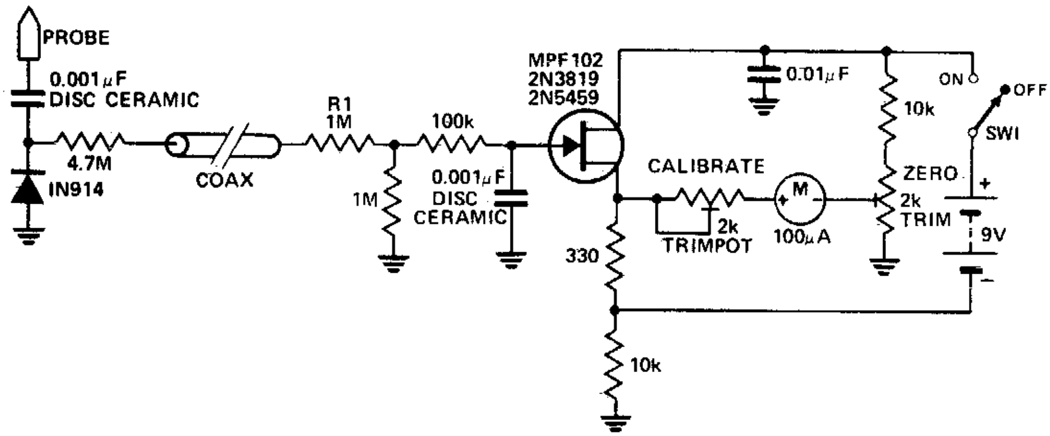


Fig. 49-15

Circuit Notes

This circuit measures RF voltages beyond 200 MHz and up to about 5 V. The diode should be mounted in a remote probe, close to the probe tip. Sensitivity is excellent and voltages less than 1 V peak can be easily measured. The unit can be calibrated by connecting the input to a known level of RF voltage, such as a calibrated signal generator, and setting the calibrate control.

MINIMUM COMPONENT TACHOMETER

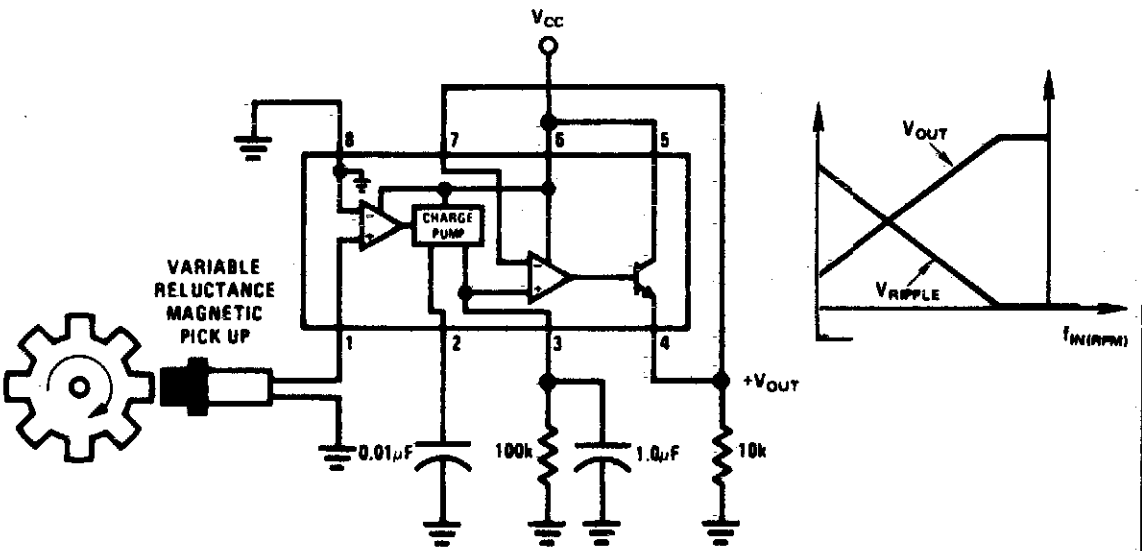
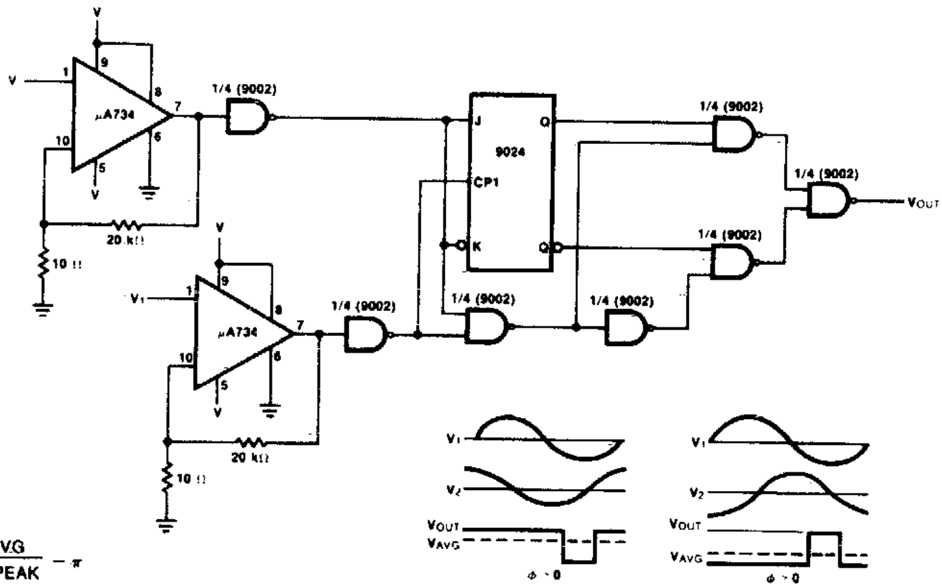


Fig. 49-16

PHASE METER



$$\phi = \frac{2\pi\, V_{AVG}}{V_{OUT,\,PEAK}} - \pi$$

Fig. 49-17

PRECISION CALIBRATION STANDARD

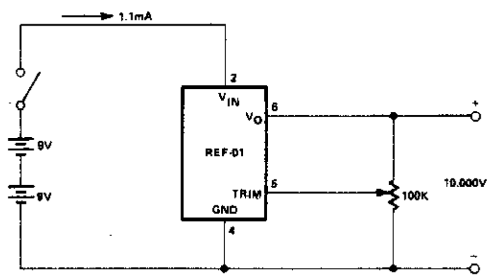


Fig. 49-18

ZENER DIODE CHECKER

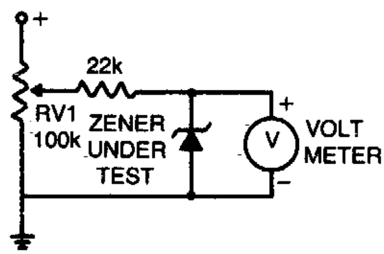


Fig. 49-19

Circuit Notes

An external power supply that gives a voltage higher than the highest expected rating of the zener diodes to be tested is required. Potentiometer RV1 is adjusted until the meter reading stabilizes. This reading is the zener diode's breakdown voltage.

50

Metal Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

MICROPOWER METAL DETECTOR

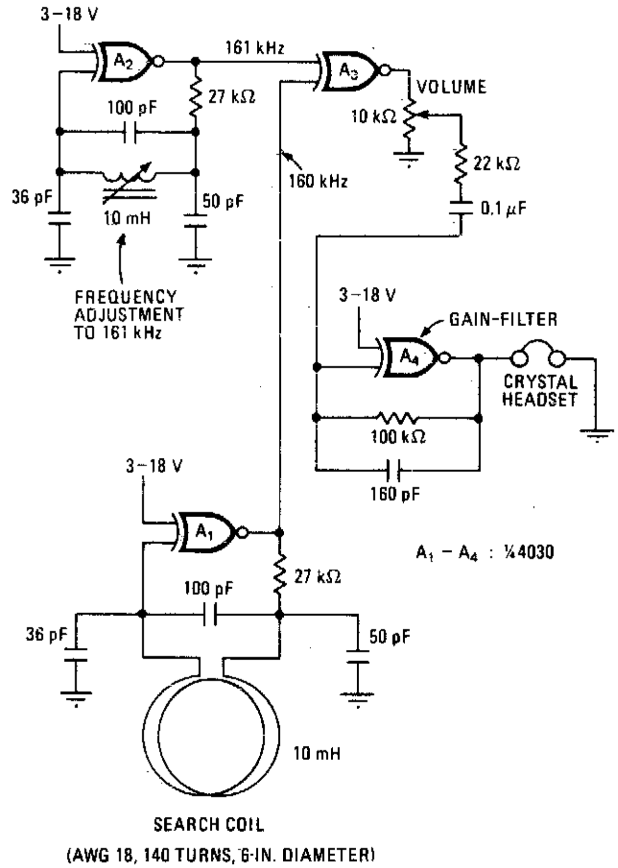


Fig. 50-1

Circuit Notes

This battery-powered metal detector uses four exclusive-OR gates contained in the 4030 CMOS integrated circuit. The gates are wired as a twin-oscillators and a search coil serves as the inductance element in one of the oscillators. When the coil is brought near metal, the resultant change in its effective inductance changes the oscillator's frequency. Gates A1 and A2 form the two oscillators which are tuned to 160 and 161 kilohertz respectively. The pulses produced by each oscillator are mixed in A3, its output contains sum and difference frequencies at 1 and 321 kHz. The 321 kHz signal is filtered out by the 10 kHz low-pass filter at A4, leaving the 1 kHz signal to be amplified for the crystal headset connected at the output. The device's sensitivity is sufficient to detect coin-sized objects a foot away.

LO-PARTS TREASURE LOCATOR

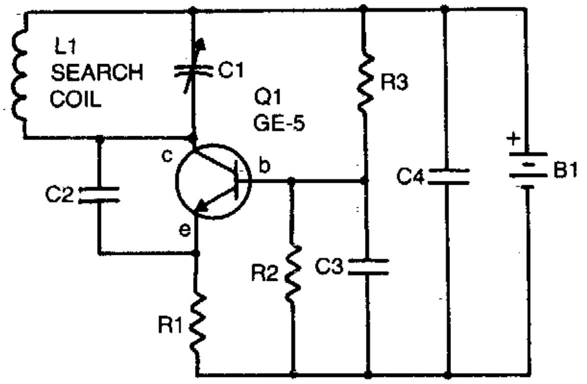


Fig. 50-2

PARTS LIST FOR
LO-PARTS TREASURE LOCATOR
B1—9-Vdc transistor battery
C1—365-pF trimmer or variable capacitor
C2—100-pF, 100-V silver mica capacitor
C3—0.05-μF, disc capacitor
C4—4.7- or 5-μF, 12-V electrolytic capacitor
L1—Search coil consisting of 18 turns of #22 enamel wire scramble wound on 4-in. diameter form
Q1—RCA SK3011 npn transistor or equiv.
R1—680-ohm, ½-watt resistor
R2—10,000-ohm, ½-watt resistor
R3—47,000-ohm, ½-watt resistor

Circuit Notes

Locator uses a transistor radio as the detector. With the radio tuned to a weak station, adjust C1 so the locator oscillator beats against the received signal. When the search head passes over metal, the inductance of L1 changes thereby changing the locator oscillator's frequency and changing the beat tone in the radio. The search coil consists of 18 turns of #22 enameled wire scramble wound on a 4-in. diameter form. After the coil is wound and checked for proper operation, saturate the coil with RTV adhesive for stable operation of the locator.

51

Metronomes

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Accentuated Beat Metronome

Sight N' Sound Metronome

Micrometronome

ACCENTUATED BEAT METRONOME

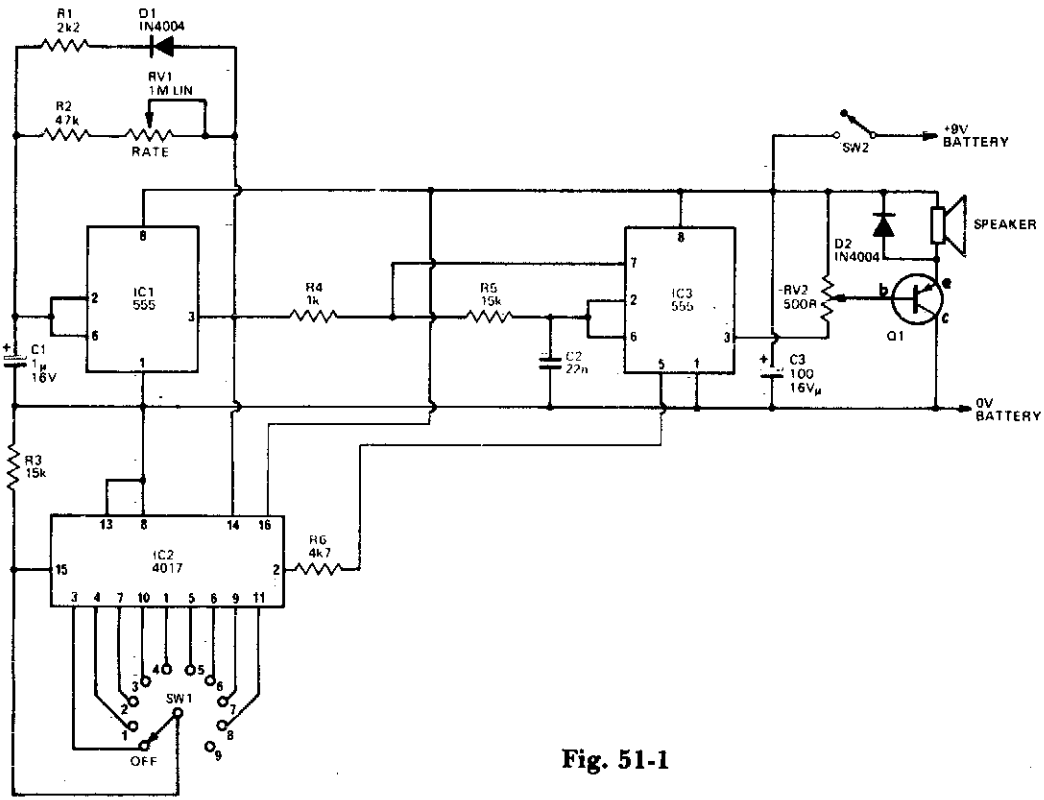


Fig. 51-1

Circuit Notes

IC3 acts as an oscillator which operates if the output of IC1 is high. With the values used the two frequencies produced are about 800 Hz and 2500 Hz. The output is buffered by Q1 which drives the speaker. The first IC is used to generate the tone duration and the time interval between beats. The interval is adjustable by RV1 while the tone duration is set by R1. The output of IC1 also clocks IC2, a decade counter with 10 decoded outputs. Each of these outputs go high in sequence on each clock. The second output of IC2 is connected to the control input of IC3 and is used to change the frequency. Therefore the first tone will be high frequency, the second low and the third to tenth will be high again. This gives the 9-1 beat. If for example the 5th output is connected to the reset, the first tone will be high, the second low, and the third and fourth high, then when the 5th output goes to a high it resets it back to the first which is a high tone. We then have 3 high and one low tones or a 3-1.

SIGHT N' SOUND METRONOME

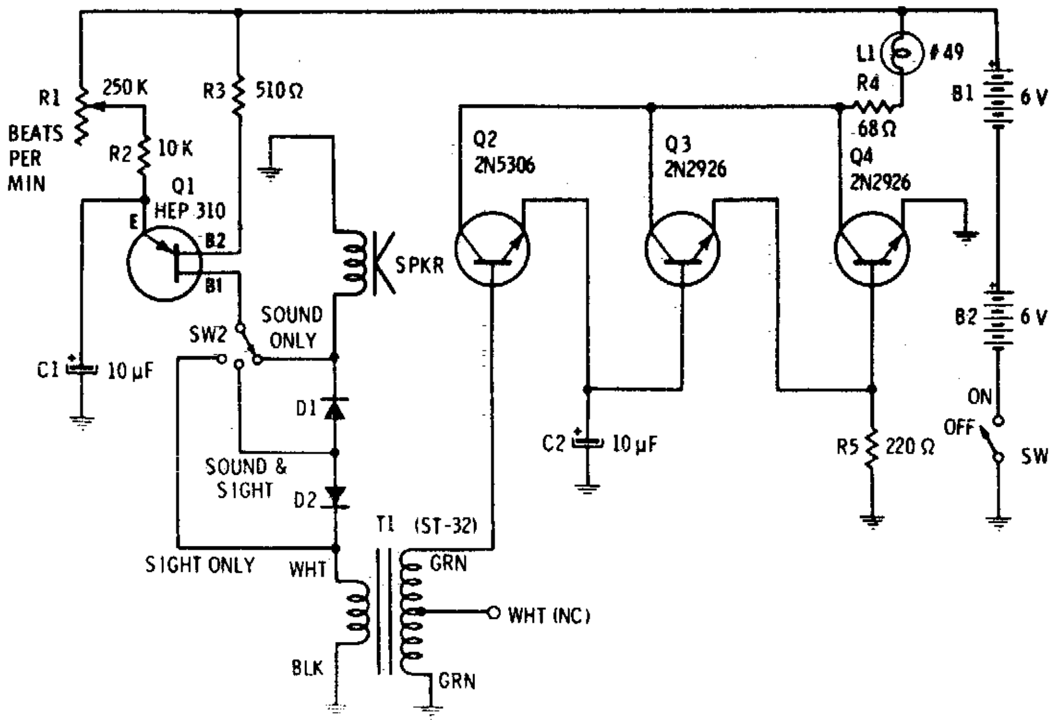


Fig. 51-2

Circuit Notes

Precise, adjustable control of beats per minute from a largo of 18 to a frenzied, high presto of 500. These beats are produced acoustically through a speaker. A light flashes at the same rate. When SW1 is closed, C1 begins to charge through R1 and R2. C1 will eventually reach a voltage at which the emitter of unijunction transistor is switched on, "dumping" the energy stored in C1 into an 8 ohm speaker. To produce a distinct "plop", brief pulses across T2 secondary drive Q2 into conduction. The extra gain of Q3 and Q4 are sufficient to briefly switch L1 on, then off, as the pulse wave passes. Capacitor C2 "stretches" the pulse slightly to overcome the thermal inertia of the lamp, so that a bright flash occurs.

MICROMETRONOME

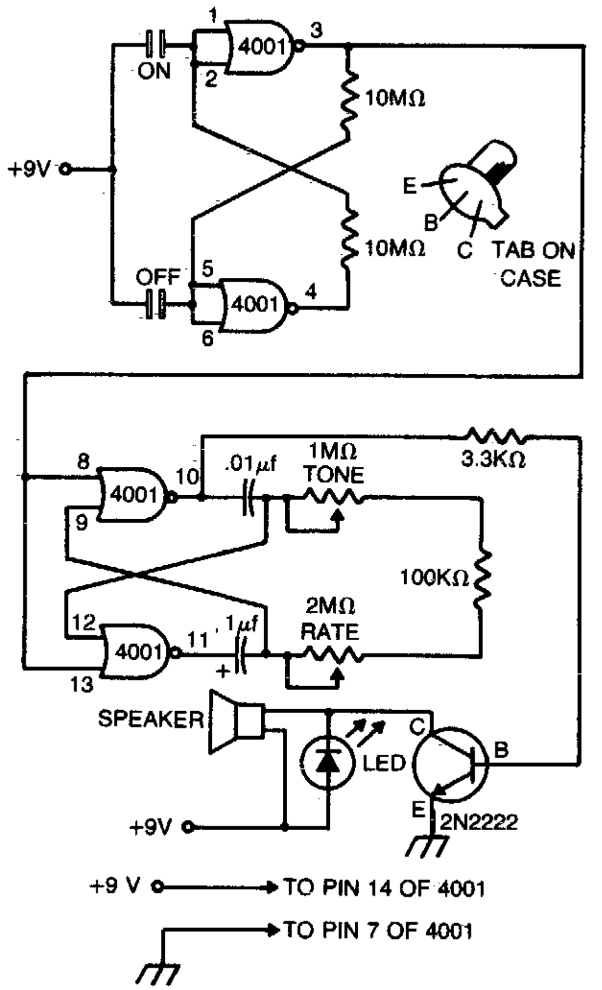


Fig. 51-3

Circuit Notes

This compact metronome will run for years on a single nine-volt transistor battery. Has both tone and pulse rate controls, and uses touch plates to start and stop, can be built in a case no larger than a pack of cigarettes. The touch plates consist of two strips of metal about 1/16-inch apart mounted on, but insulated from, the case. Bridging the gap closes the switch.

52

Miscellaneous Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

INTERCOM

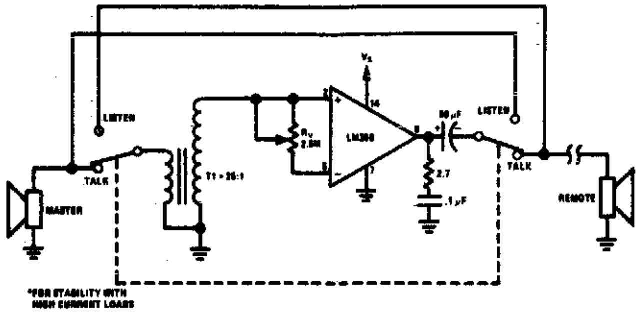


Fig. 52-1

Circuit Notes

The circuit provides a minimum component intercom. With switch S1 in the talk position, the speaker of the master station acts as the microphone with the aid of step-up transformer T1. A turns ratio of 25 and a device gain of 50 allows a maximum loop gain of 1250. R_v provides a common mode volume control. Switching S1 to the listen position reverses the role of the master and remote speakers.

MUSICAL ORGAN

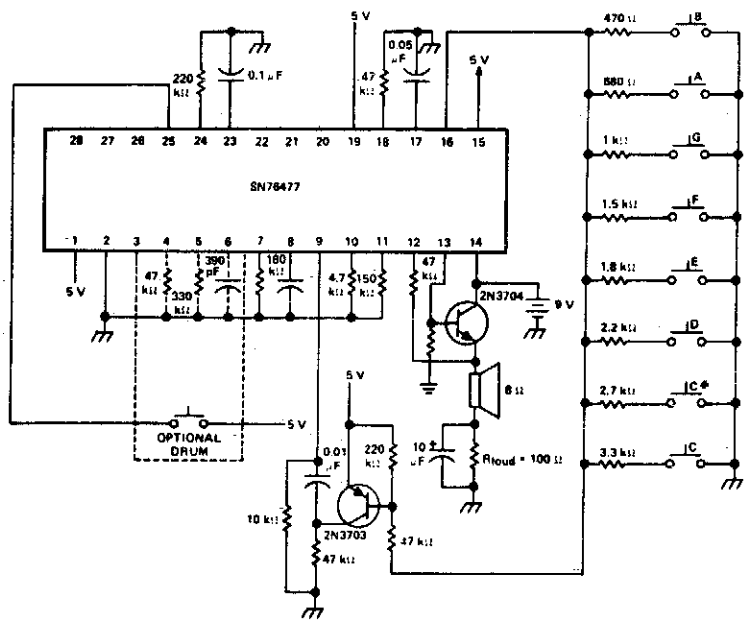


Fig. 52-2

LASER DIODE PULSER

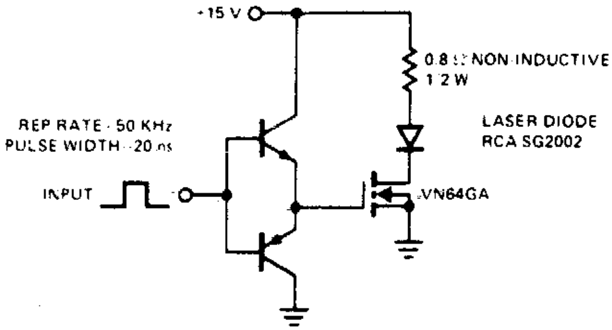
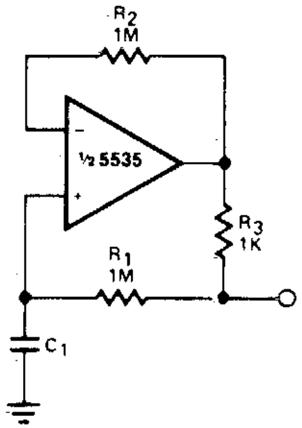


Fig. 52-3

Circuit Notes

This drive is capable of driving the laser-diode with 10 ampere, 20 ns pulses. For a 0.1% duty cycle, the repetition rate will be 50 kHz. A complementary emitter-follower is used as a driver. Switching speed is determined by the f_T of the bipolar transistors used and the impedance of the drive source.

CAPACITANCE MULTIPLIER



$$C_{eff} = \frac{R_1}{R_3} \cdot C_1$$

$$R_S = R_3$$

Fig. 52-4

All resistor values are in ohms

Circuit Notes

This circuit can be used to simulate large capacitances using small value components. With the values shown and C = 10 μF, an effective capacitance of 10,000 μF was obtained. The Q available is limited by the effective series resistance. So R1 should be as large as practical.

SIMULATED INDUCTOR

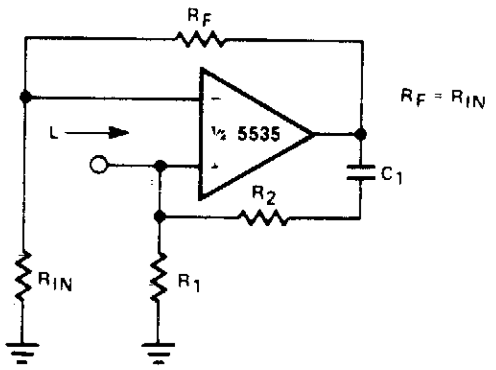


Fig. 52-5

Circuit Notes

With a constant current excitation, the voltage dropped across an inductance increases with frequency. Thus, an active device whose output increases with frequency can be characterized as an inductance. The circuit yields such a response with the effective inductance being equal to: L = R1R2C. The Q of this inductance depends upon R1 being equal to R2. At the same time, however, the positive and negative feedback paths of the amplifier are equal leading to the distinct possibility of instability at high frequencies. R1 should, therefore, always be slightly smaller than R2 to assure stable operation.

ACTIVE INDUCTOR

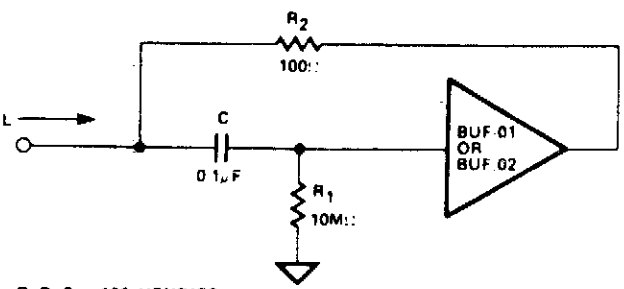


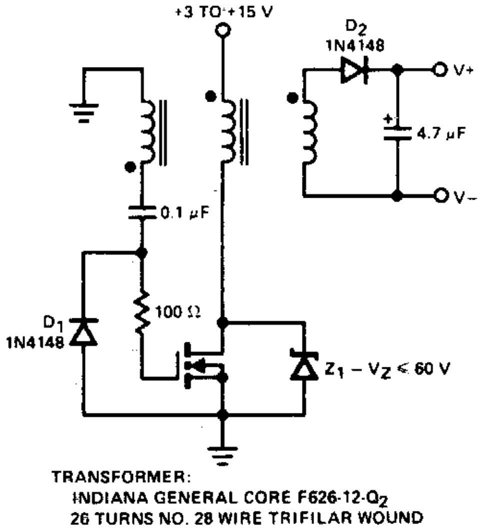
Fig. 52-6

$L \geq R_1R_2C = 100$ HENRIES
$R_S = R_2 = 100\Omega$
$R_P = R_1 = 10$ MEGΩ
ASSUMING C_{STRAY} (ACROSS R_1) OF 5 pF THE UPPER FREQUENCY LIMIT IS APPROXIMATELY 7kHz.
$X_L = 100\Omega$ AT $f = 0.159$Hz

Circuit Notes

An active inductor is realized with an eight-lead IC, two carbon resistors, and a small capacitor. A commercial inductor of 50 henries may occupy up to five cubic inches.

POSITIVE INPUT/NEGATIVE OUTPUT CHARGE PUMP



Circuit Notes

A simple means of generating a low-power voltage supply of opposite polarity from the main supply. Self oscillating driver produces pulses at a repetition frequency of 100 kHz. When the VMOS device is off, capacitor C is charged to the positive supply. When the VMOS transistor switches on, C delivers a negative voltage through the series diode to the output. The zener serves as a dissipative regulator.

Fig. 52-7

SHIFT REGISTER DRIVER

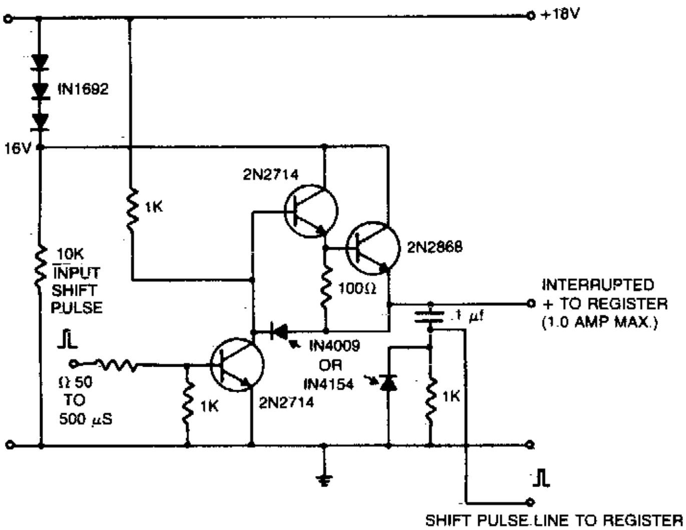


Fig. 52-8

Circuit Notes

A 16 V power supply can be synthesized as shown using IN1692 rectifiers. A shift pulse input saturates the 2N2714 depriving the Darlington combination (2N2714 and 2N2868) of base drive. The negative pulse so generated on the 15 V line is differentiated to produce a positive trigger pulse at its trailing edge.

TAPE RECORDER

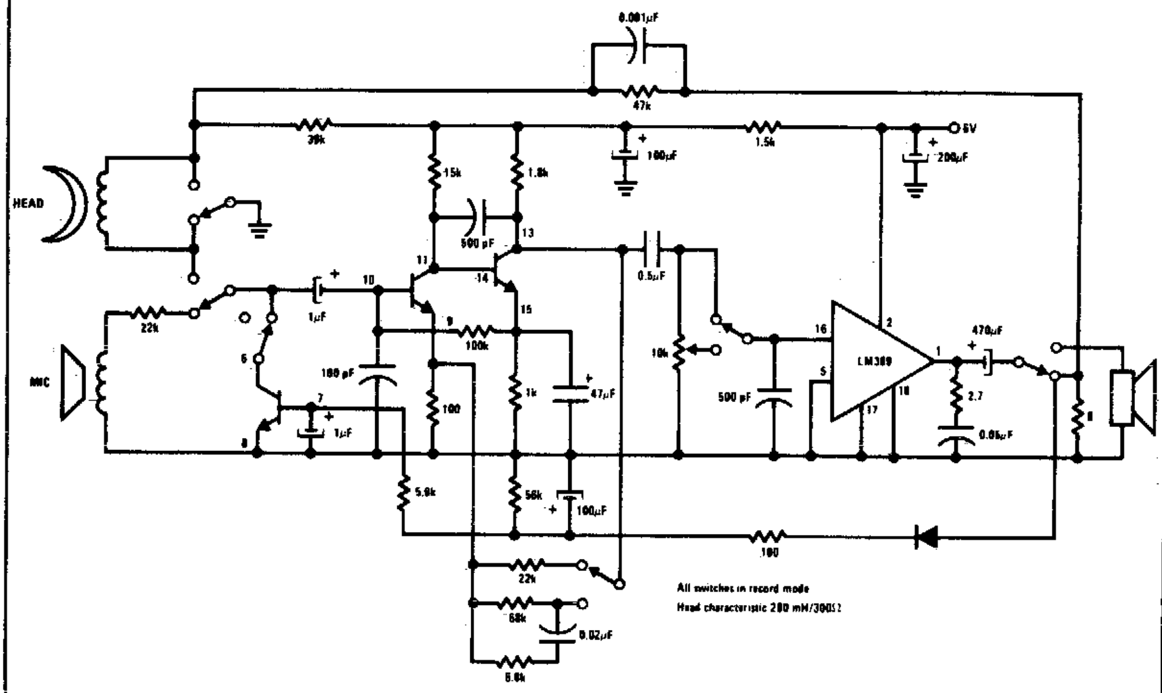


Fig. 52-9

Circuit Notes

Complete record/playback cassette tape machine amplifier. Two of the transistors act as signal amplifiers, with the third used for automatic level control during the record mode.

NEGATIVE-EDGE DIFFERENTIATOR

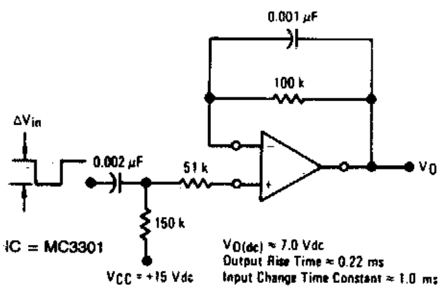


Fig. 52-10

STYLUS ORGAN

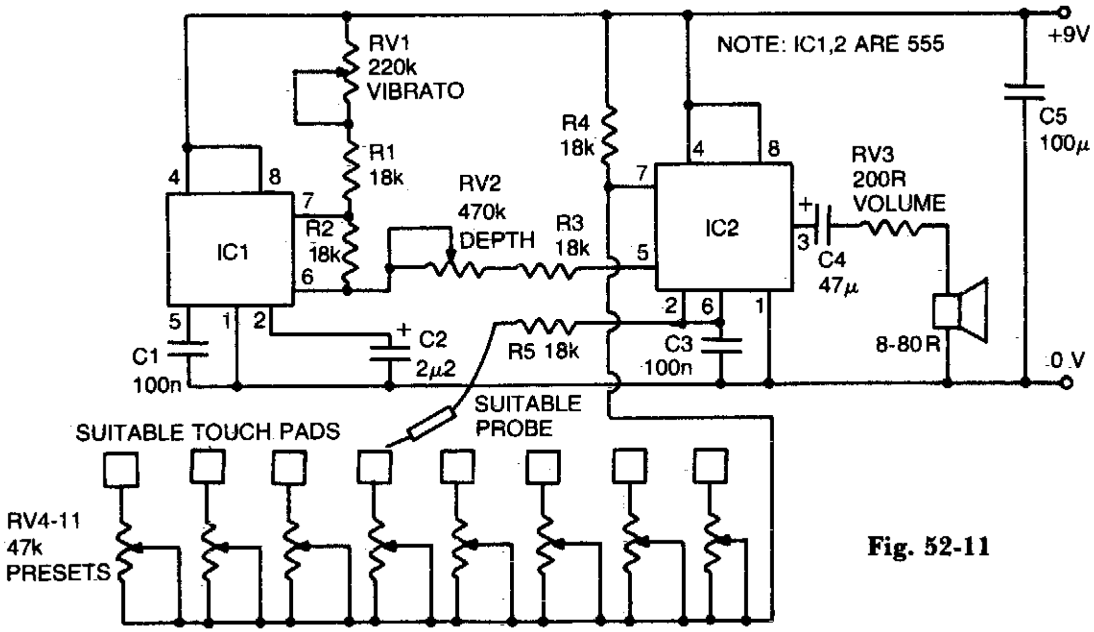


Fig. 52-11

Circuit Notes

IC2 is an audio frequency oscillator. Its frequency is primarily controlled by the resistance between pins 2 and 7. RV4-11 control the oscillator frequency and by touching a stylus (connected via limiting resistor R5 to pin 2) to each preset, different notes can be played. IC1 is a low frequency oscillator (approximately 3-10Hz), the frequency of which is variable by RV1. The output of this oscillator is connected through depth control RV2 and limiting resistor R3 to the voltage control input of the audio frequency oscillator. Thus a vibrato effect occurs.

POSITIVE-EDGE DIFFERENTIATOR

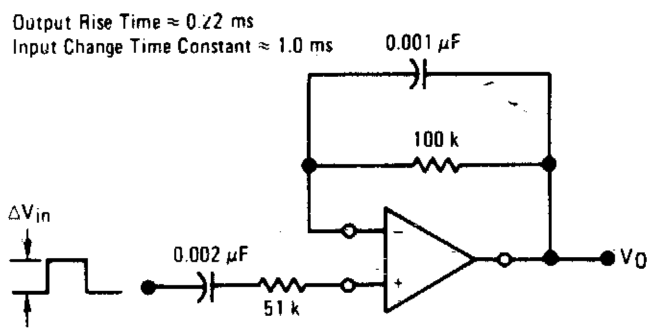


Fig. 52-12

FOUR CHANNEL DATA ACQUISITION SYSTEM

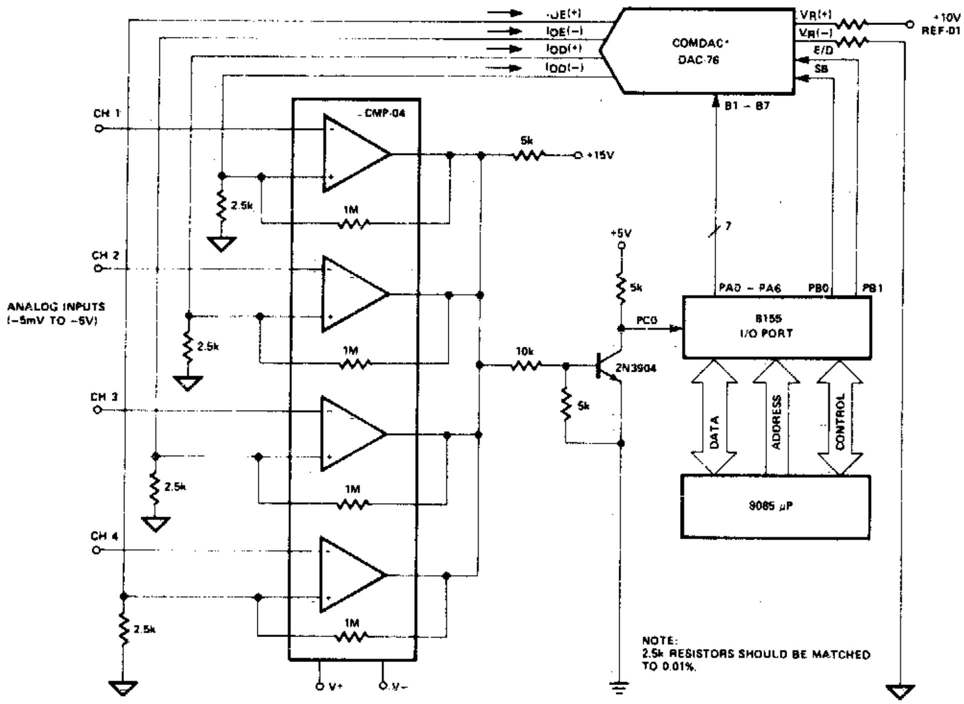


Fig. 52-13

TRIAC TRIGGER

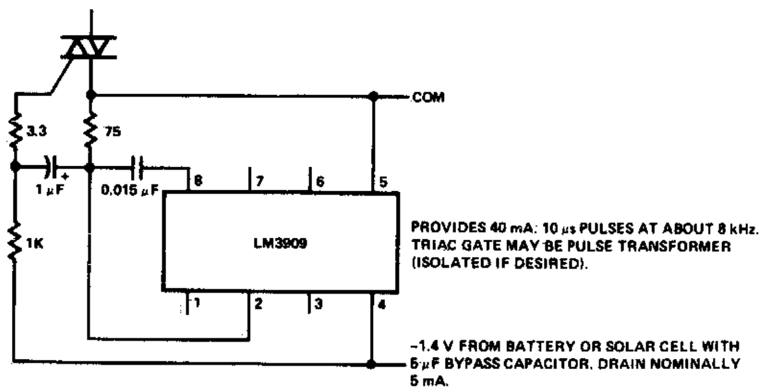


Fig. 52-14

PRECISION RECTIFIERS

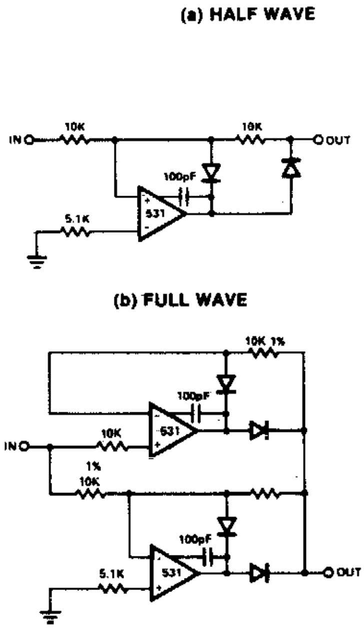


Fig. 52-15

FAST INVERTER CIRCUIT

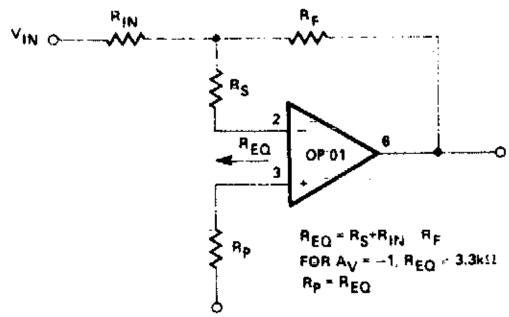


Fig. 52-17

VOLTAGE CONTROL RESISTOR

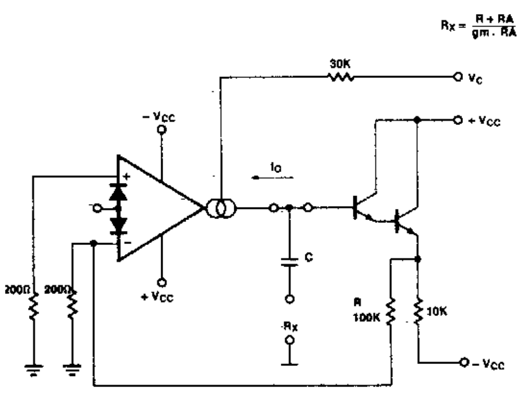


Fig. 52-16

INVERSE SCALER

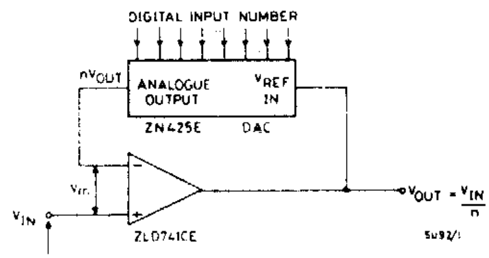


Fig. 52-18

Circuit Notes

If a DAC is operated in the feedback loop of an operational amplifier, then the amplifier gain is inversely proportional to the input digital number or code to the DAC. The version giving scaling inversely proportional to positive voltage is shown.

5.0 V SQUARE WAVE CALIBRATOR

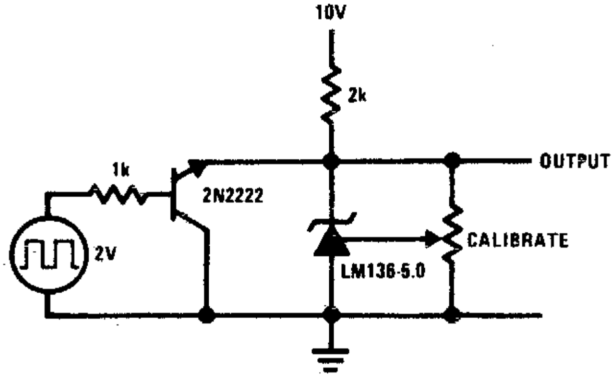


Fig. 52-19

LOW DRIFT INTEGRATOR AND LOW-LEAKAGE GUARDED RESET

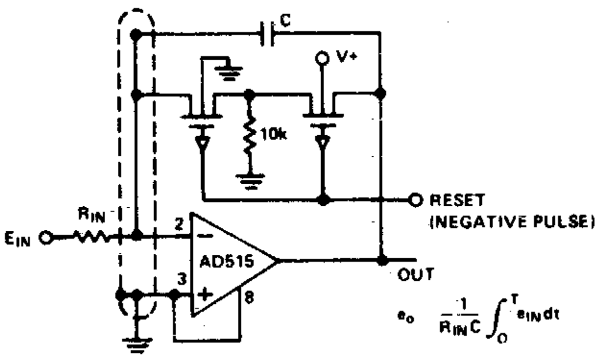
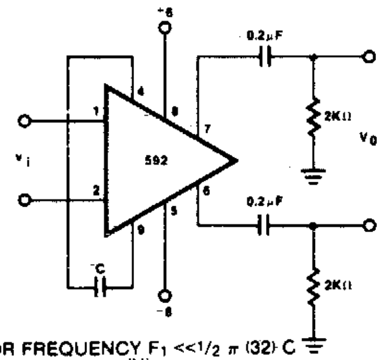


Fig. 52-20

DIFFERENTIATOR WITH HIGH COMMON MODE NOISE REJECTION



FOR FREQUENCY $F_1 << 1/2\ \pi\ (32)\ C$

$V_0 \approx 1.4 \times 10^4 C \frac{dVi}{dT}$

Fig. 52-21

DIGITAL TRANSMISSION ISOLATOR

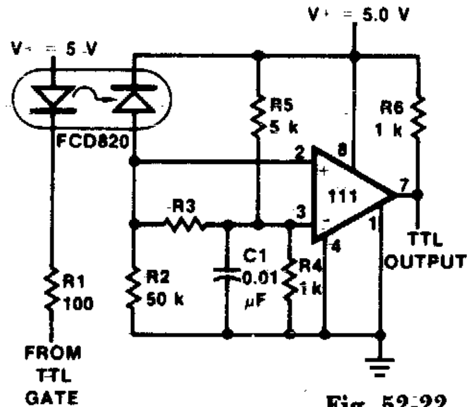


Fig. 52-22

53

Mixers and Multiplexers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DIFFERENTIAL MUX/DEMUX SYSTEM

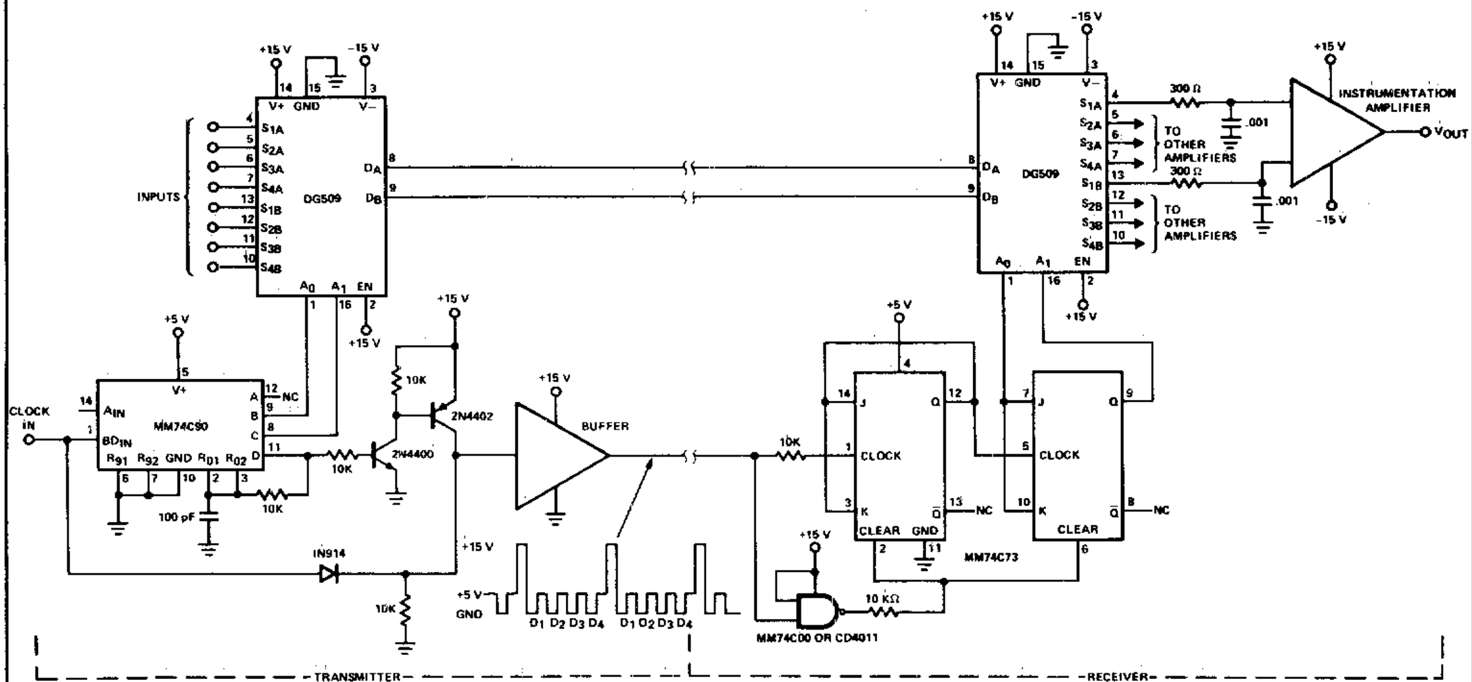


Fig. 53-1

EIGHT CHANNEL MUX/DEMUX SYSTEM

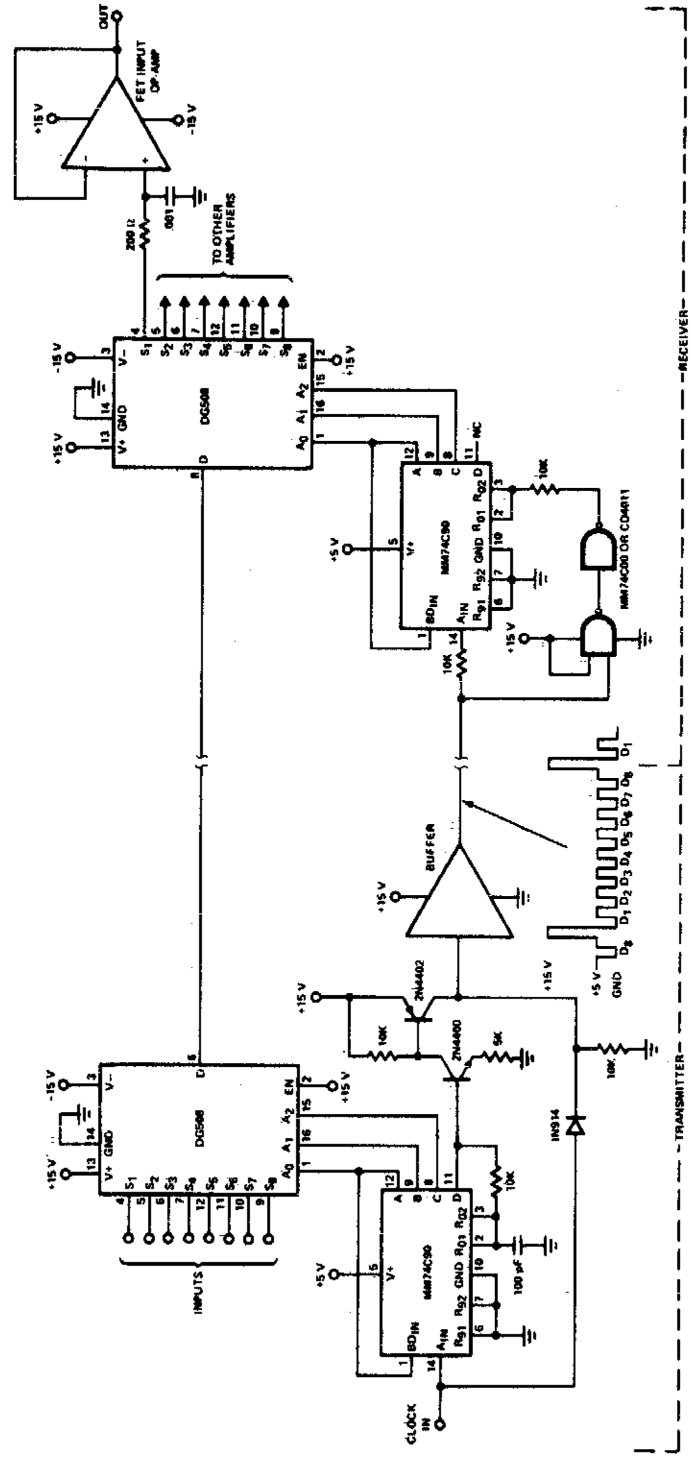
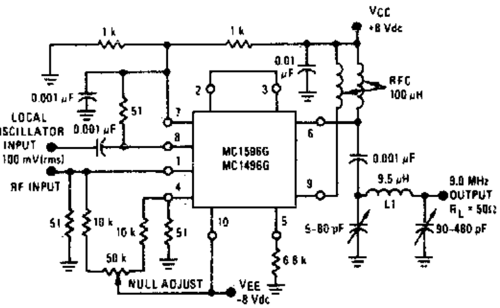


Fig. 53-2

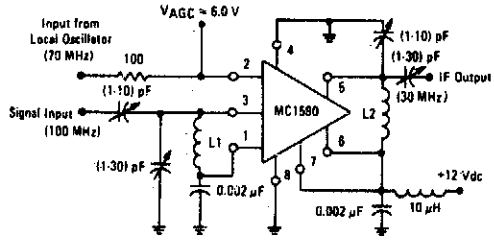
DOUBLY BALANCED MIXER (BROADBAND INPUTS, 9.0 MHz TUNED OUTPUT)



L1 = 44 TURNS AWG NO. 28 ENAMELED WIRE, WOUND ON MICROMETALS TYPE 44-6 TOROID CORE.

Fig. 53-3

100 MHz MIXER

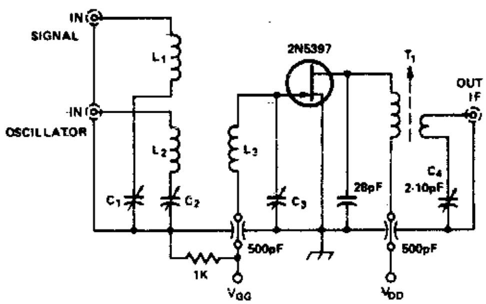


L1 = 5 Turns, #16 AWG Wire, 1/4" ID, 5/8" Long

L2 = 16 Turns, #20 AWG Wire on a Toroid Core, (T44-6 Micro Metal or Equiv)

Fig. 53-5

COMMON-SOURCE MIXER



L_1 – 1.4" long; #22 enamel, close coupled to L_3

L_2 – 1.6" long; #22 enamel, close coupled to L_3

L_3 – 1.75" long; #18 copper

T_1 – Primary, 13T; #22 enamel, close wound on 1/4" form ≈ 1μH

Secondary, 3T; #22 enamel, close wound over primary

C_1, C_2, C_3, C_4, – 0.8 to 12pF; Johanson type 2950

Fig. 53-4

MULTIPLEXER/MIXER

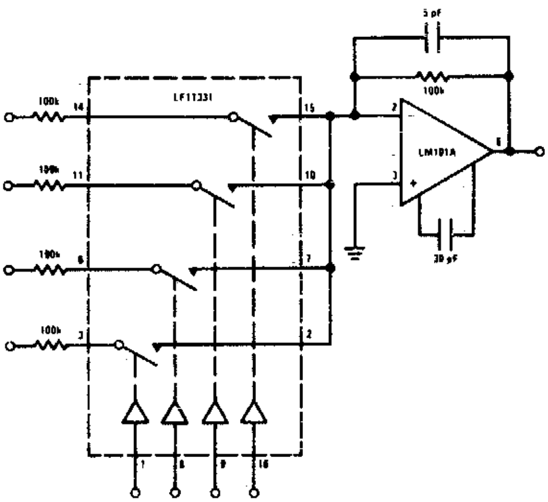


Fig. 53-6

WIDE BAND DIFFERENTIAL MULTIPLEXER

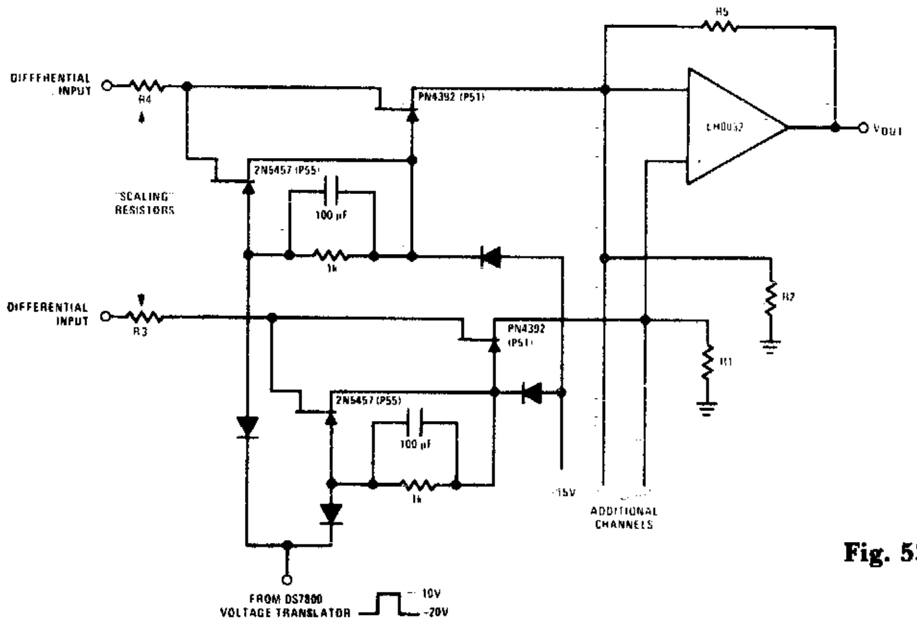


Fig. 53-7

Circuit Notes

This design allows high frequency signal handling and high toggle rates simultaneously. Toggle rates up to 1 MHz and MHz signals are possible with this circuit.

54

Modulation Monitors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Modulation Monitor   Visual Modulation Indicator

CB Modulation Monitor

MODULATION MONITOR

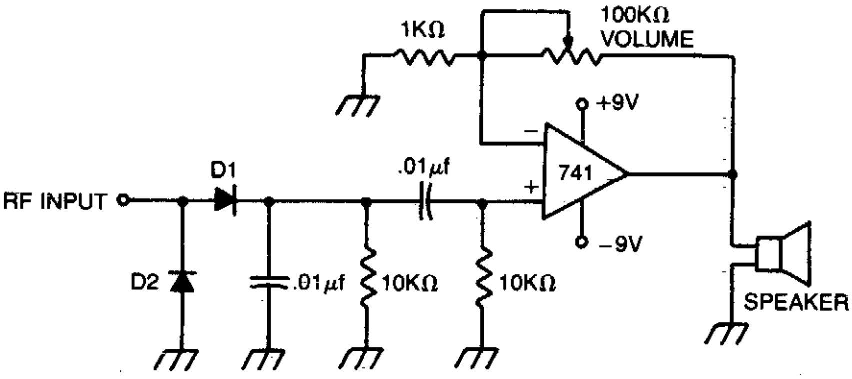


Fig. 54-1

Circuit Notes

Broad-tuned receiver demodulates the RF signal picked up by a loosely coupled wire placed near the transmitting antenna.

VISUAL MODULATION INDICATOR

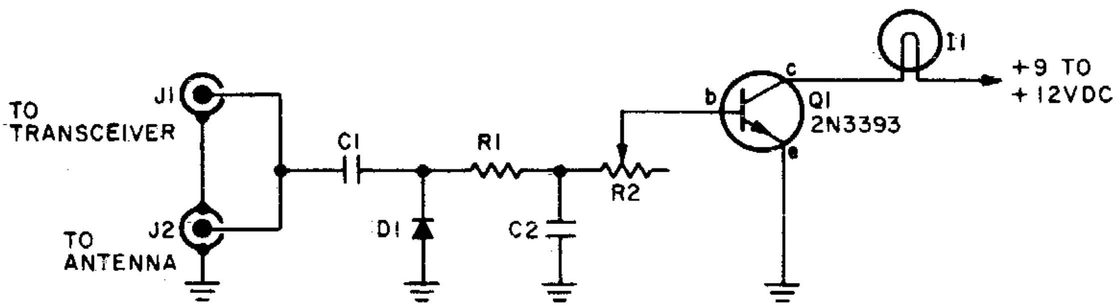


Fig. 54-2

Circuit Notes

Indicator lamp brightness varies in step with modulated RF signal. Adjust R2 with transmitter on (modulated) until the lamp flashes in step with modulation. C1 = 5 pf, C2 = 100 pF, D1 = 1N60 or 1N34 (Germanium), R3 = 10 K pot, I1 = 6-8 V, 30-60 mA incandescent bulb, Q1 = 2N3393 (for increased sensitivity use 2N3392 or other high-gain transistor).

CB MODULATION MONITOR

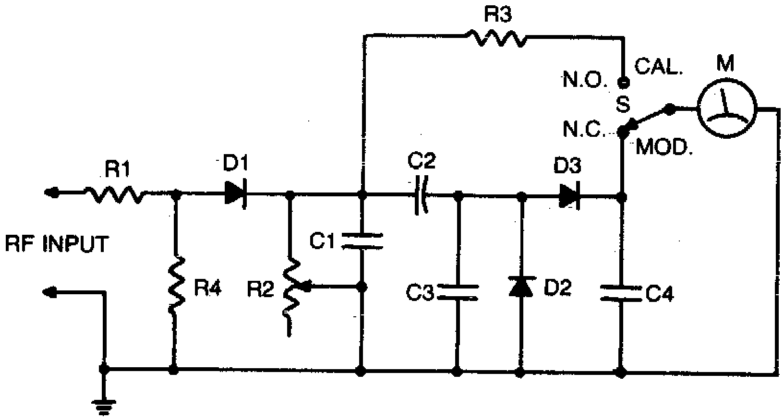


Fig. 54-3

PARTS LIST

C1—500-pF, 100-Vdc capacitor
C2—10-μF, 10-Vdc electrolytic capacitor
C3—200-pF, 100-Vdc capacitor
C4—300-pF, 100-Vdc capacitor
D1, D2, D3—1N60
M1—0-1 mA DC high-speed meter
R1, R4—1000-ohm, ½-watt resistor
R2—1000-ohm pot
R3—910-ohm, ½-watt resistor, 5%
S1—Spdt spring-return switch

Circuit Notes

Connect this circuit to a transceiver with a coaxial T connector in the transmission line. Key the transmitter (unmodulated), set S1 to CAL, and adjust R2 for a full scale reading. Return S1 to MOD position. The meter will read % modulation with 10% accuracy.

55

Modulators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TV MODULATOR

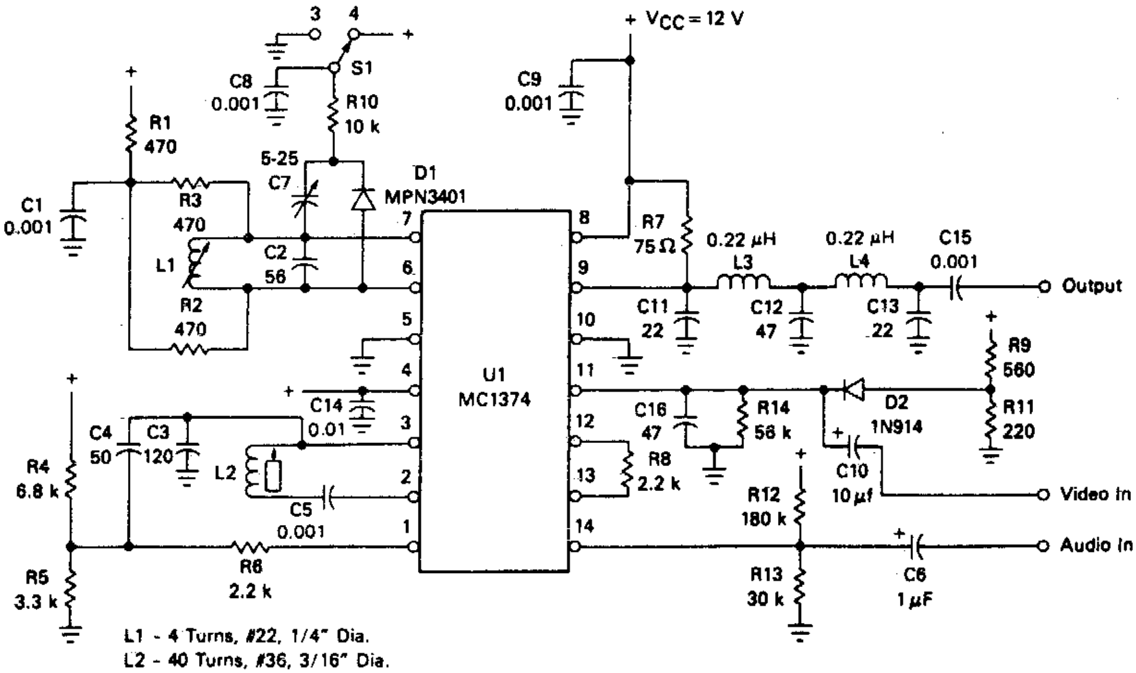


Fig. 55-1

Circuit Notes

The FM oscillator/modulator is a voltage-controlled oscillator, which exhibits a nearly linear output frequency versus input voltage characteristic for a wide deviation. It provides a good FM source with a few inexpensive external parts. It has a frequency range of 1.4 to 14 MHz and can typically produce a ±25 kHz modulated 4.5 MHz signal with about 0.6% total harmonic distortion.

TV MODULATOR

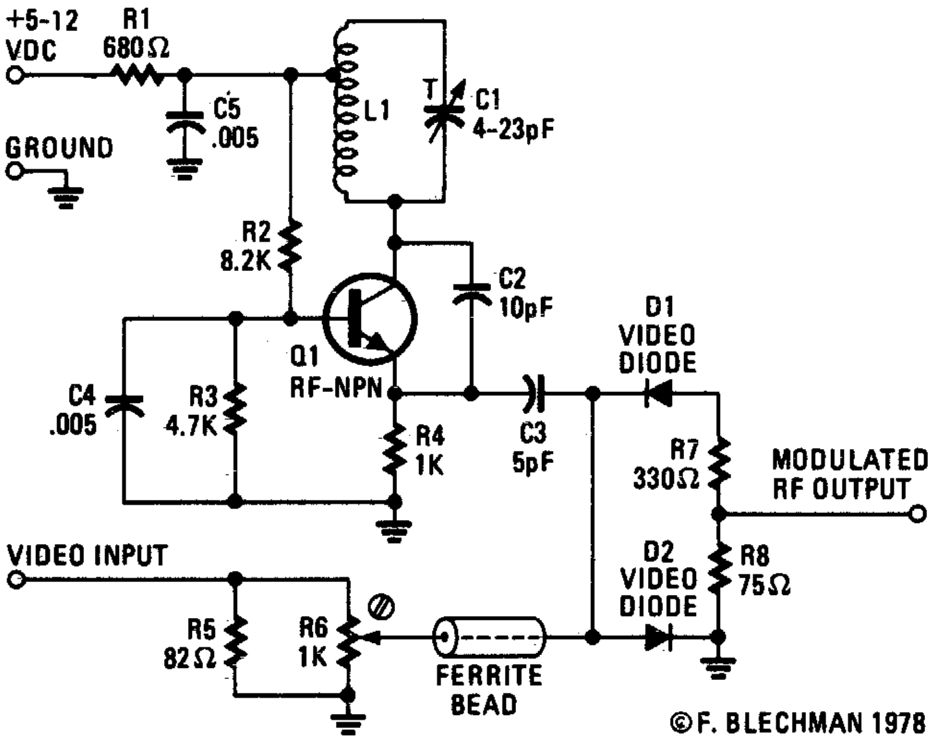
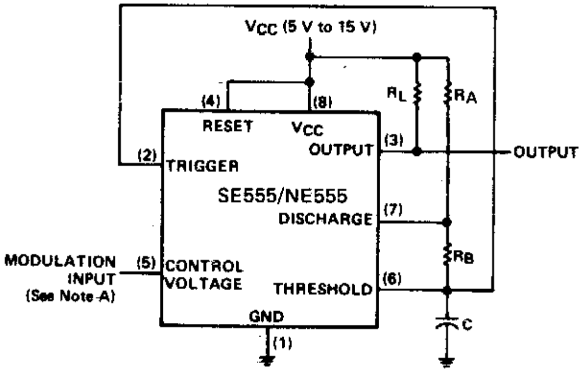


Fig. 55-2

Circuit Notes

The VHF frequency is generated by a tuned Hartley oscillator circuit. Resistors R2, R3, and R4 bias the transistor, with tapped inductor L1 and trimmer capacitor C1 forming the tank circuit. Adjusting C1 determines the frequency. Capacitor C2 provides positive feedback from the tank circuit to the emitter at Q1. Capacitor C4 provides an RF ground for the base of Q1. Bypass capacitor C5 and resistor R1 filter out the radio frequencies generated in the tank circuit to prevent radiation from the power-supply lines. The video signal enters the parallel combination of resistors R5 and R6; this combination closely matches the 75 ohm impedance of most video cables. Resistor R6 is a small screwdriver-adjusted potentiometer that is used to control the video input level to mixer diodes D1 and D2.

PULSE-POSITION MODULATOR



NOTE A: The modulating signal may be direct or capacitively coupled to the control voltage terminal. For direct coupling, the effects of modulation source voltage and impedance on the bias of the SE555/NE555 should be considered.

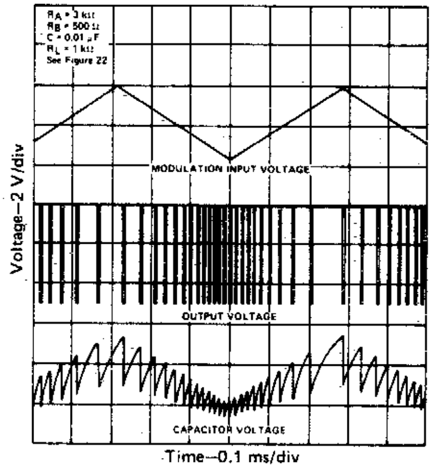


Fig. 55-3

Circuit Notes

The threshold voltage, and thereby the time delay, of a free-running oscillator is shown modulated with a triangular-wave modulation signal; however, any modulating wave-shape could be used.

PULSE-WIDTH MODULATOR

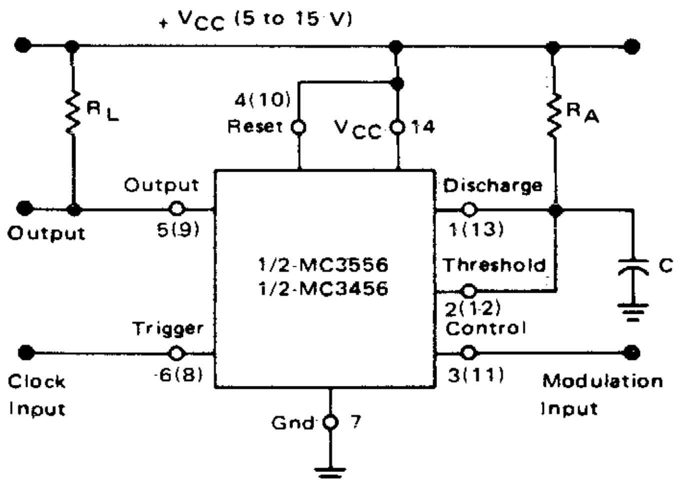
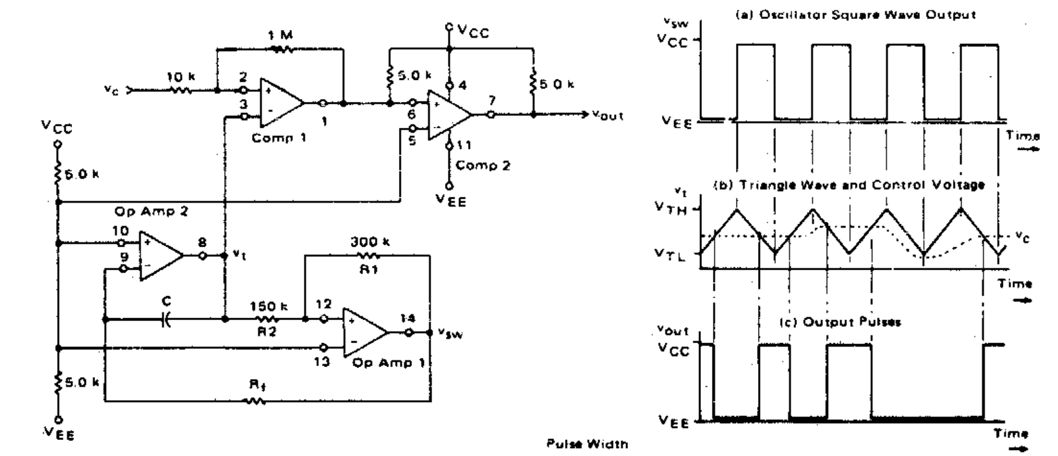


Fig. 55-4

Circuit Notes

If the timer is triggered with a continuous pulse train in the monostable mode of operation, the charge time of the capacitor can be varied by changing the control voltage at pin 3. In this manner, the output pulse width can be modulated by applying a modulating signal that controls the threshold voltage.

PULSE-WIDTH MODULATOR



$V_{TH} = \frac{1}{2}V_S(1 + R2/R1) + V_{EE}$ $\quad V_S = V_{CC} - V_{EE}$

$V_{TL} = \frac{1}{2}V_S(1 - R2/R1) + V_{EE}$

Oscillator Frequency

$f = \frac{R1}{4R_fCR2}$

IC = MC3405

Pulse Width

$P.W. = \left(\frac{1}{f}\right)\left(\frac{V_c - V_{TL}}{V_{TH} - V_{TL}}\right)$ When: $V_{TL} < V_C < V_{TH}$

Duty Cycle in %

$D.C. = \left(\frac{V_c - V_{TL}}{V_{TH} - V_{TL}}\right)(100)$

Fig. 55-5

RF MODULATOR

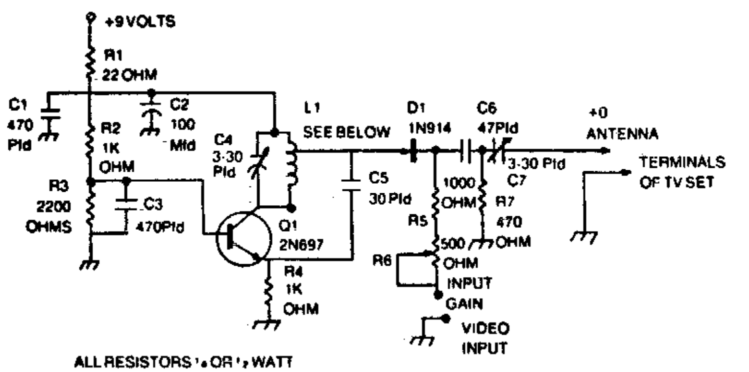


Fig. 55-6

Circuit Notes

Capacitors C1, C3, C5, and C6 should be dipped mica. C4 and C7 are compression or piston trimmer types. R6 is PC-board mount trimpot. L1 is 6 turns of No. 14 enameled wire, ⅜ inch I.D. by ¾ inch long, tapped at 1 turn from top.

LINEAR PULSE-WIDTH MODULATOR

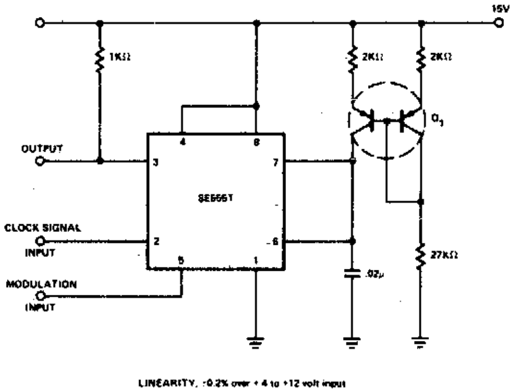


Fig. 55-7

VIDEO MODULATOR

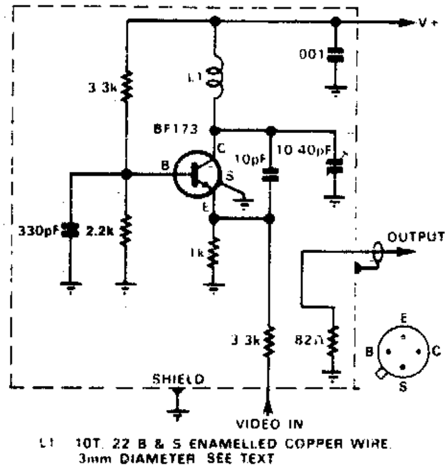


Fig. 55-9

BALANCED MODULATOR (+12 Vdc SINGLE SUPPLY)

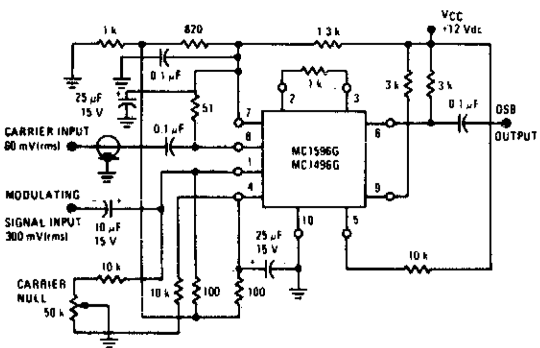


Fig. 55-8

MODULATOR

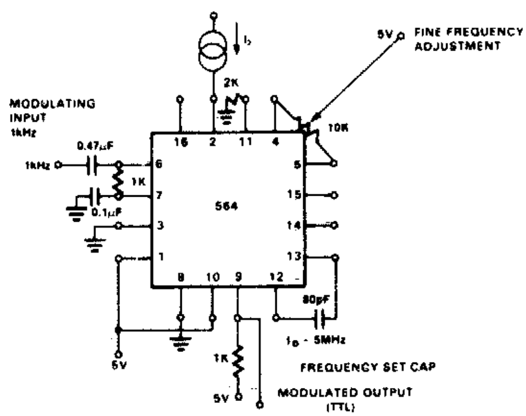
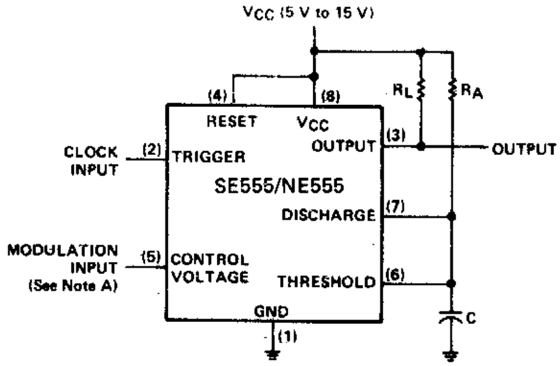


Fig. 55-10

PULSE-WIDTH MODULATOR



NOTE A: The modulating signal may be direct or capacitively coupled to the control voltage terminal. For direct coupling, the effects of modulation source voltage and impedance on the bias of the SE555/NE555 should be considered.

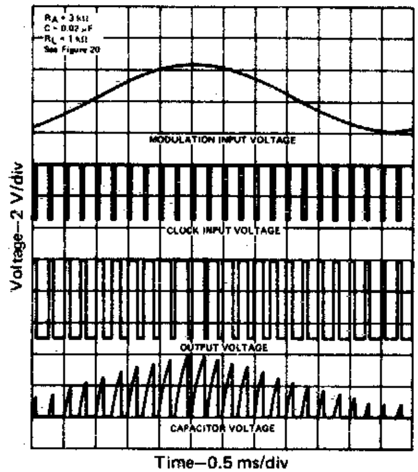


Fig. 55-11

Circuit Notes

The monostable circuit is triggered by a continuous input pulse train and the threshold voltage is modulated by a control signal. The resultant effect is a modulation of the output pulse width, as shown. A sine-wave modulation signal is illustrated, but any wave-shape could be used.

AM MODULATOR

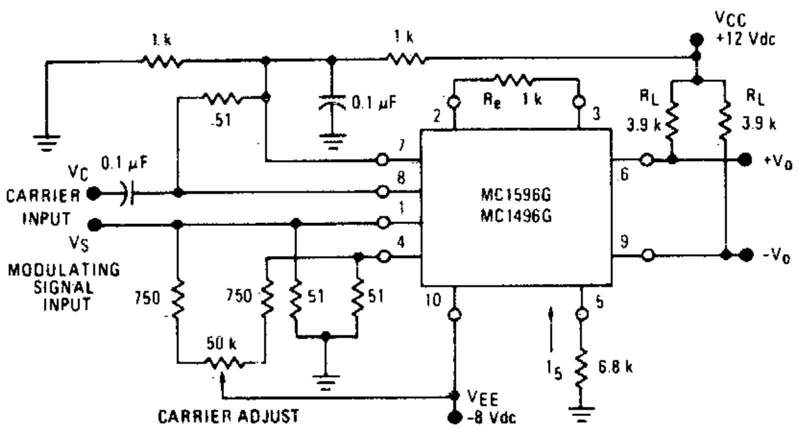


Fig. 55-12

TV MODULATOR USING A MOTOROLA MC1374

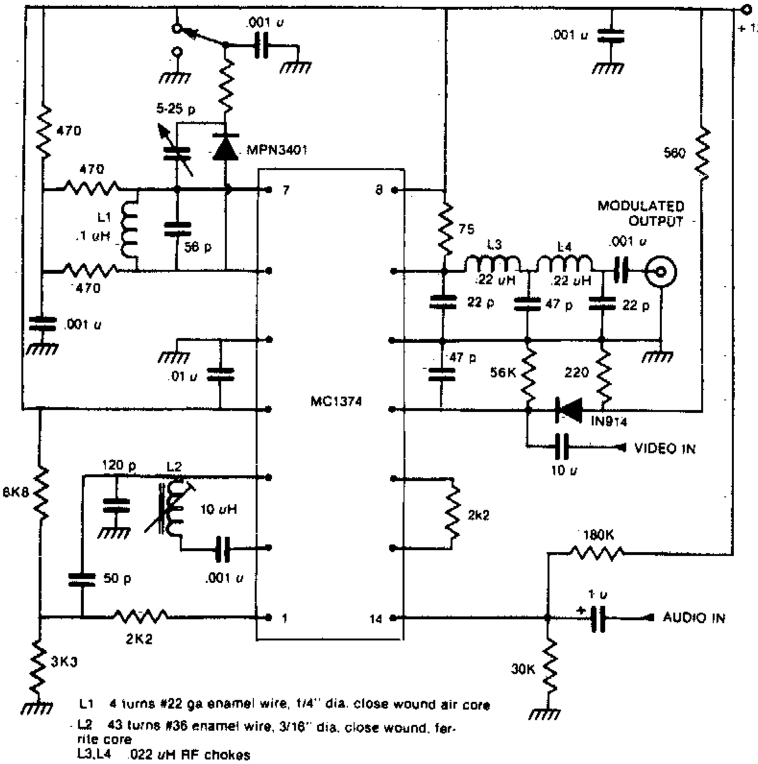
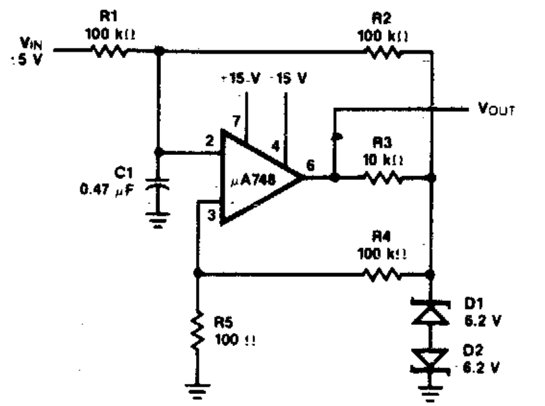


Fig. 55-13

Circuit Notes

This one-chip modulator requires some outboard circuitry and a shielded box.

PULSE-WIDTH MODULATOR



$$f_c = \frac{1}{2\pi R2\, C1}$$

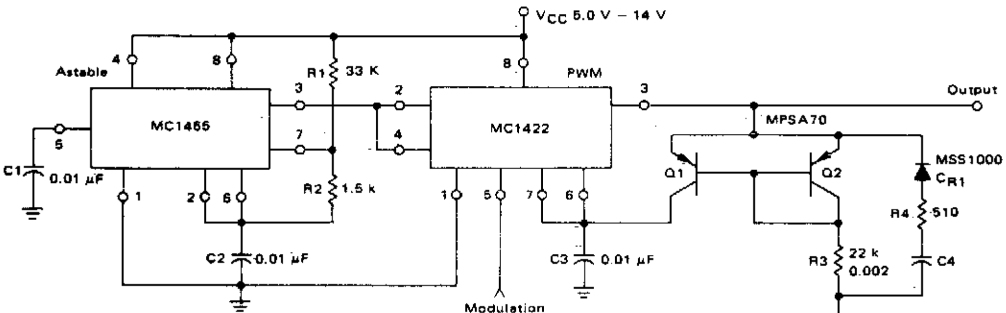
$$f_n = \frac{1}{2\pi R1\, C1}$$

$$= \frac{1}{2\pi R2\, C2}$$

$$f_c < f_n < f_{unity\ gain}$$

Fig. 55-14

PULSE-WIDTH MODULATOR



Note 1. MC1422 can be utilized as an astable if an external 2/3 ratio resistive divider is used at pin 5.

2. See waveforms

– PULSE WIDTH MODULATOR WAVEFORMS

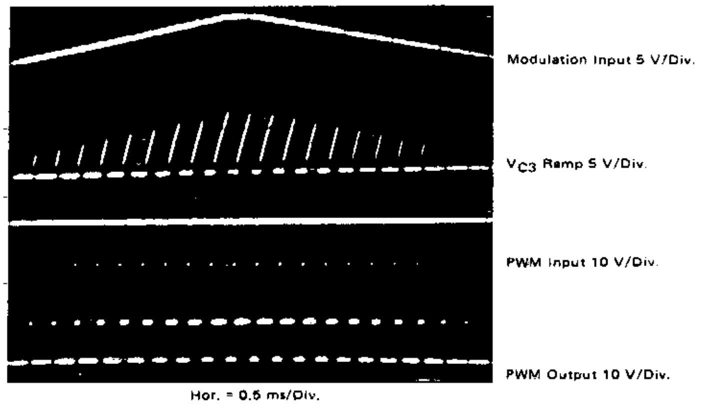


Fig. 55-15

VHF MODULATOR

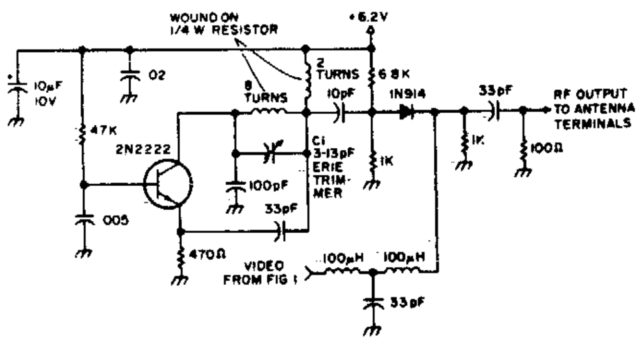


Fig. 55-16

56

Moisture and Rain Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RAIN ALARM

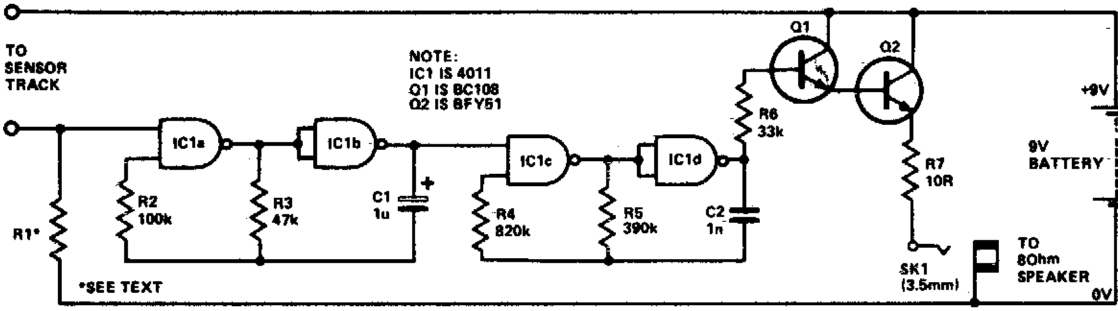
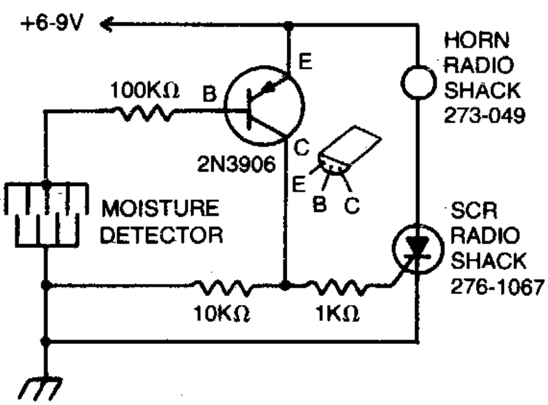


Fig. 56-1

Circuit Notes

The circuit uses four NAND gates of a 4011 package. In each oscillator, while one gate is configured as a straightforward inverter, the other has one input that can act as a control input. Oscillator action is inhibited if this input is held low. The first oscillator (IC1a and IC1b) has this input tied low via a high value resistor (R1) that acts as a sensitivity control. Thus this oscillator will be disabled until the control input is taken high. Any moisture bridging the sensor track will so enable the output which is a square wave at about 10 Hz. This in turn will gate on and off the 500 Hz oscillator formed by IC1c and IC1d. This latter oscillator drives the loudspeaker via R6, the Darlington pair formed by Q1 and Q2, and resistor R7.

MOISTURE DETECTOR



Circuit Notes

The detector is made of fine wires spaced about one or two inches apart. When the area between a pair of wires becomes moistened, the horn will sound. To turn it off, dc power must be disconnected.

Fig. 56-2

AUTOMATIC PLANT WATERER

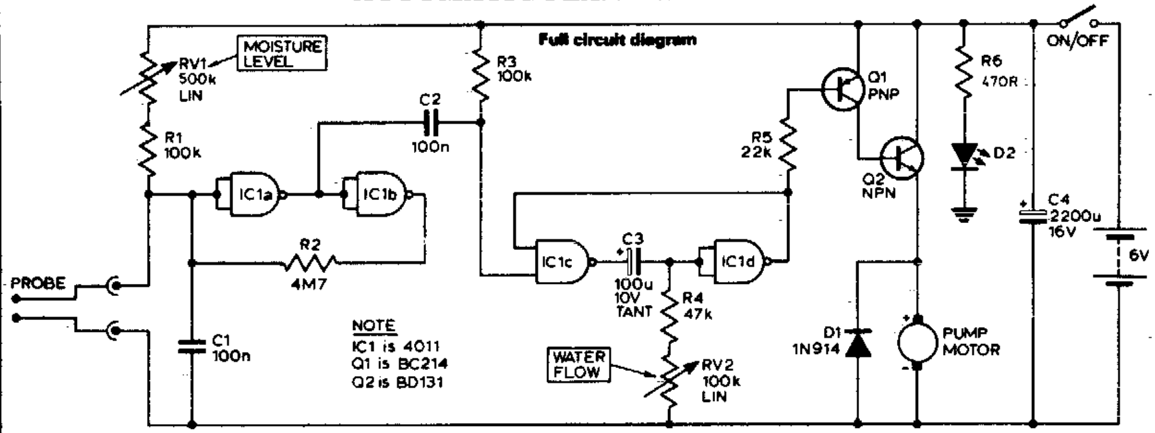


Fig. 56-3

Circuit Notes

The unit consists of a sensor, timer, and electric water pump. The sensor is embedded in the soil, and when dry, the electronics operate the water pump for a preset time. The circuit is composed of a level sensitive Schmitt trigger, variable time monostable, and output driver. When the resistance across the probe increases beyond a set value (i.e., the soil dries), the Schmitt is triggered. C2 feeds a negative going pulse to the monostable when the Schmitt triggers and R2 acts as feedback, to ensure a fast switching action.

RAIN ALARM/DOOR BELL

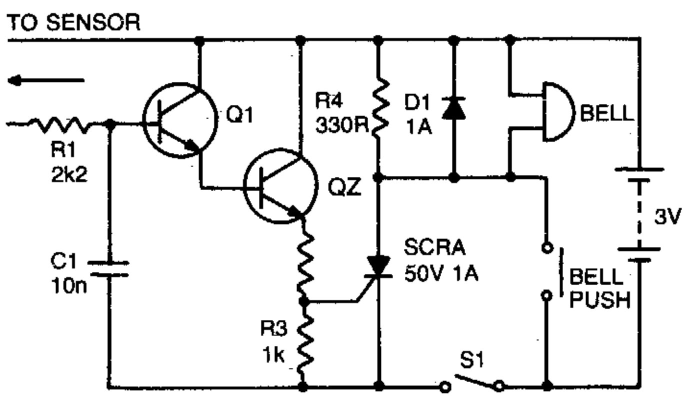


Fig. 56-4

NOTE
Q1, Q2 are 2N3706
D1 is 1N4001

Circuit Notes

With S1 open the circuit functions as a doorbell. With S1 closed, rain falling on the sensor will turn on Q1, triggering Q2 and the thyristor and activating the bell, R4 provides the holding for the thyristor while D1 prevents any damage to the thyristor from back EMF in the bell coil. The sensor can be made from 3 square inches of copper clad board with a razor cut down the center. C1 prevents any mains pickup in the sensor leads.

57

Motor Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

MOTOR SPEED CONTROL

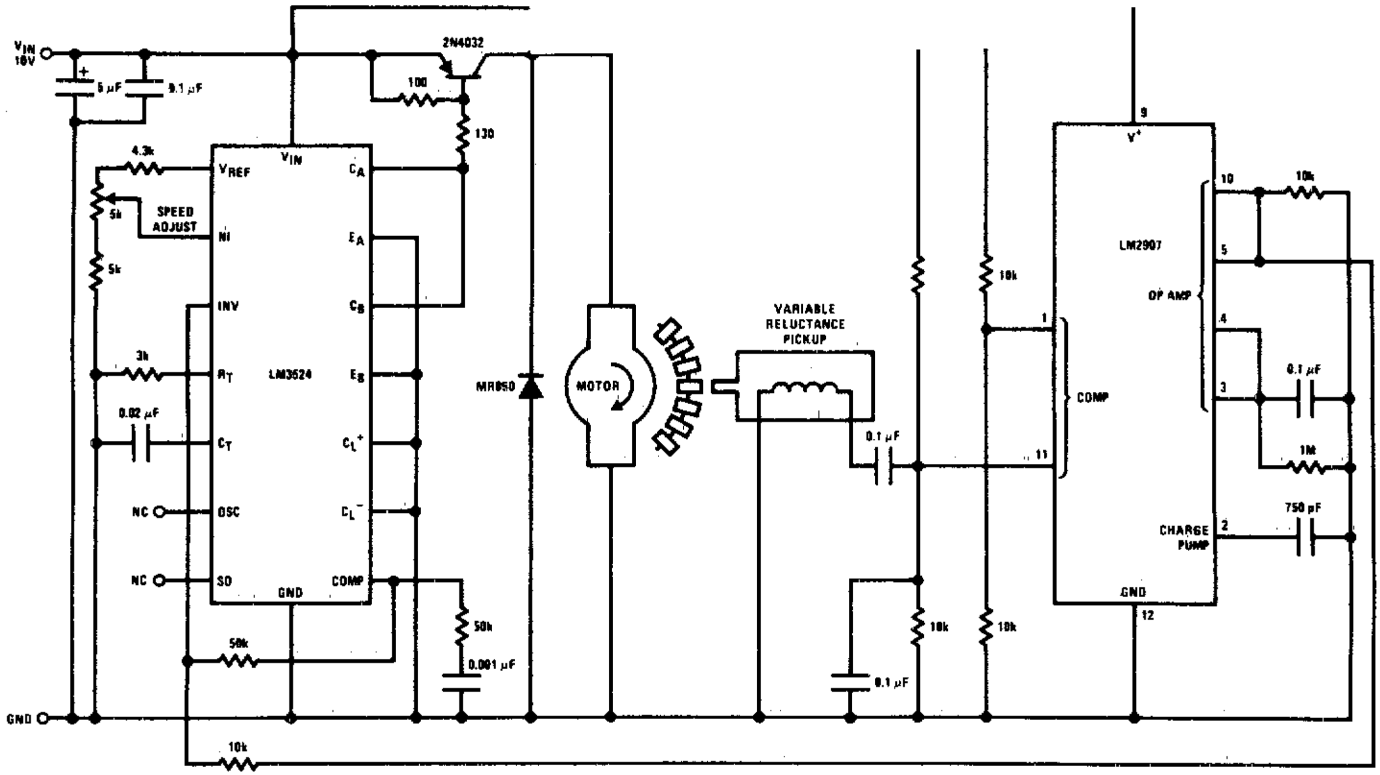
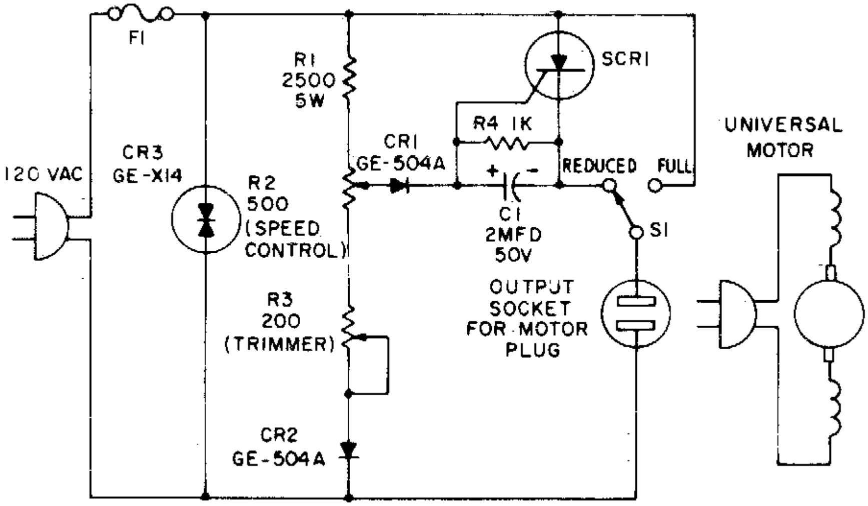


Fig. 57-1

Circuit Notes

This circuit is a regulating series dc motor speed control using the LM3524 for the control and drive for the motor and the LM2907 as a speed sensor for the feedback network.

PLUG-IN SPEED CONTROL FOR TOOLS OR APPLIANCES



COMPONENT	MOTOR NAMEPLATE RATING	
	LIGHT DUTY 3 AMP MAX	HEAVIER DUTY 5 AMP MAX
SCRI	GE-XI	GE-C30B
FI	3 AMP	5 AMP

Fig. 57-2

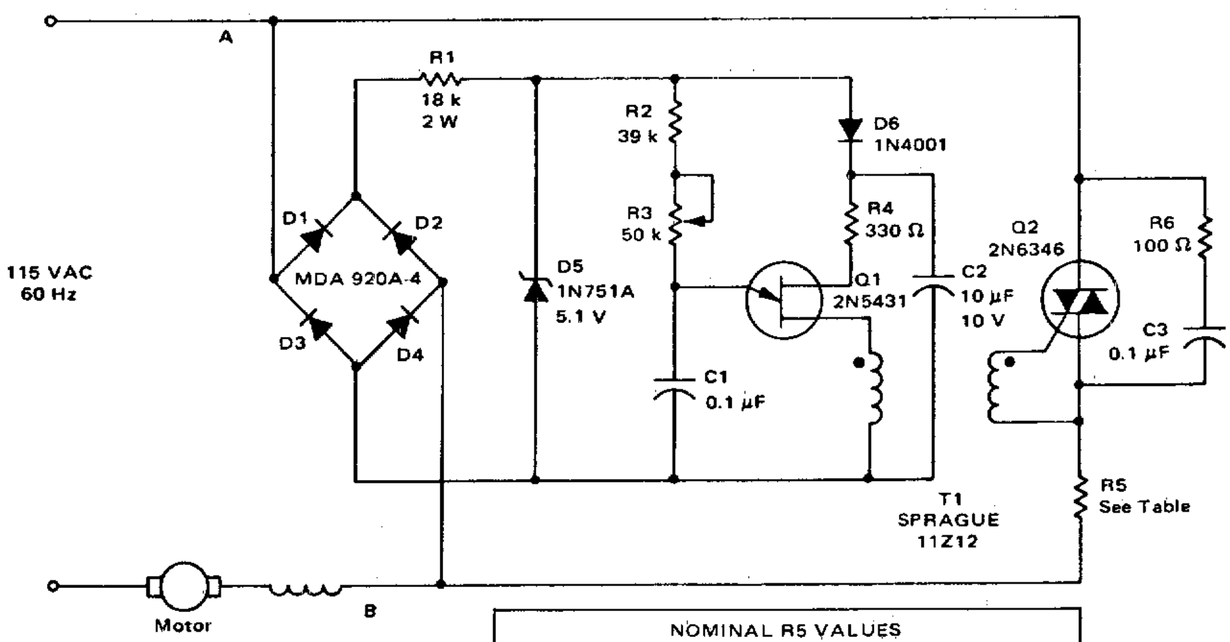
Circuit Notes

Most standard household appliances and portable hand tools can be adapted to variable-speed operation by use of this simple half-wave SCR phase control. It can be used as the speed control unit for the following typical loads provide they use series universal (brush type) motors.

Drills	Fans
Sewing Machines	Lathes
Saber saws	Vibrators
Portable band saws	Movie projectors
Food mixers	Sanders
Food blenders	

During the positive half cycle of the supply voltage, the arm on potentiometer R2 taps off a traction of the sine wave supply voltage and compares it with the counter emf of the motor through the gate of the SCR. When the pot voltage rises above the armature voltage, current flows through CR1 into the gate of the SCR, triggering it, and thus applying the remainder of that half cycle supply voltage to the motor. The speed at which the motor operates can be selected by R2. Stable operation is possible over approximately a 3-to-1 speed range.

MOTOR SPEED CONTROL WITH FEEDBACK



NOMINAL R5 VALUES			
Motor Rating (Amperes)	R5 OHMS	R5 Watts	$R5 = \frac{2}{I_M}$ I_M = Max. Rated Motor Current (RMS)
2	1	5	
3	0.67	10	
6.5	0.32	15	

Fig. 57-3

DIRECTION AND SPEED CONTROL FOR SERIES-WOUND MOTORS

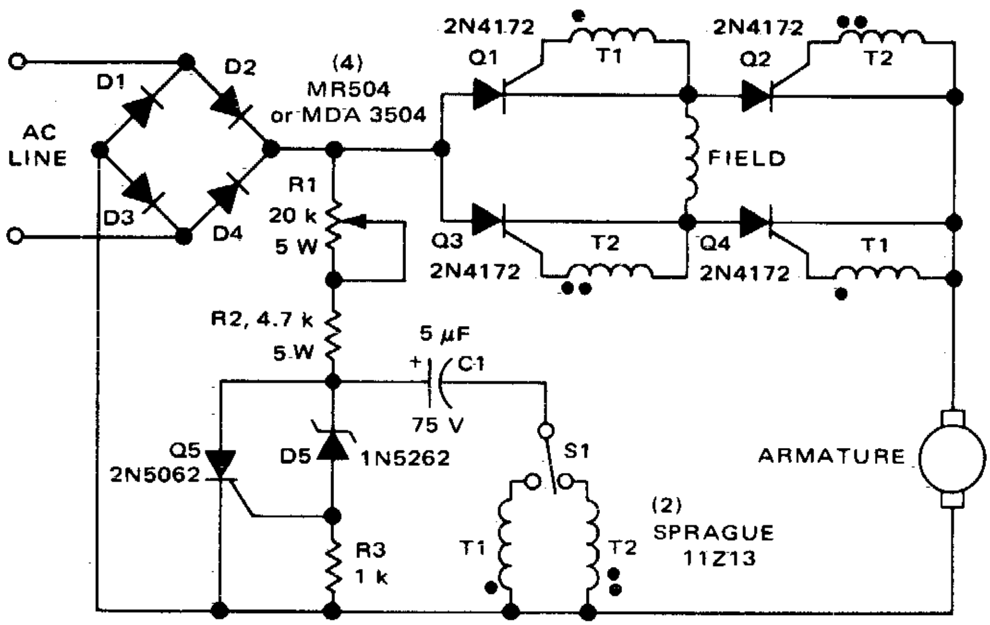
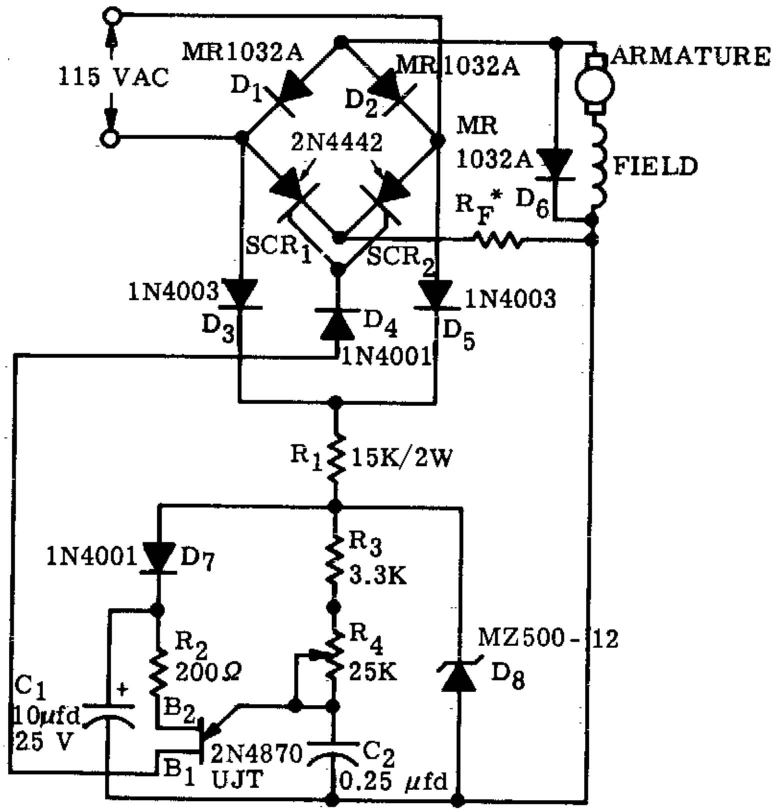


Fig. 57-4

Circuit Notes

The circuit shown here can be used to control the speed and direction of rotation of a series-wound dc motor. Silicon controlled rectifiers Q1-Q4, which are connected in a bridge arrangement, are triggered in diagonal pairs. Which pair is turned on is controlled by switch S1 since it connects either coupling transformer T1 or coupling transformer T2 to a pulsing circuit. The current in the field can be reversed by selecting either SCRs Q2 and Q3 for conduction, or SCRs Q1 and Q4 for conduction. Since the armature current is always in the same direction, the field current reverses in relation to the armature current, thus reversing the direction of rotation of the motor. A pulse circuit is used to drive the SCRs through either transformer T1 or T2. The pulse required to fire the SCR is obtained from the energy stored in capacitor C1.

HIGH-TORQUE MOTOR SPEED CONTROL



R_F* = FEEDBACK RESISTOR
(SEE TEXT)

Fig. 57-5

Circuit Notes

A bridge circuit consisting of two SCRs and two silicon rectifiers furnishes full-wave power to the motor. Diodes, D3 and D5, supply dc to the trigger circuit through dropping resistors, R1. Phase delay of SCR firing is obtained by charging C2 through resistors R3 and R4 from the voltage level established by the zener diode, D8. When C2 charges to the firing voltage of the unijunction transistor, the UJT fires, triggering the SCR that has a positive voltage on its anode. When C2 discharges sufficiently, the unijunction transistor drops out of conduction. The value of R_F is dependent upon the size of the motor and on the amount of feedback desired. A typical value for R_F can be calculated from: $R_F = \frac{2}{I_M}$ where S_{IM} is the max rated load current (rms).

MOTOR SPEED CONTROL

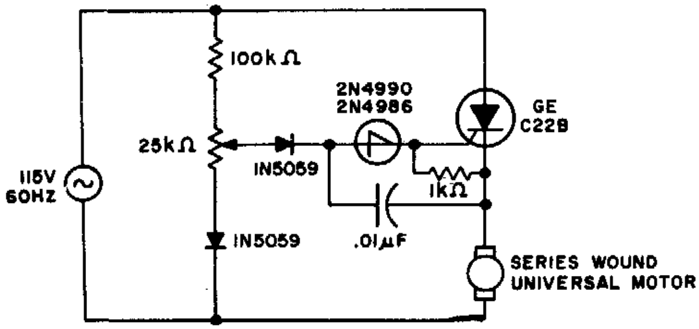
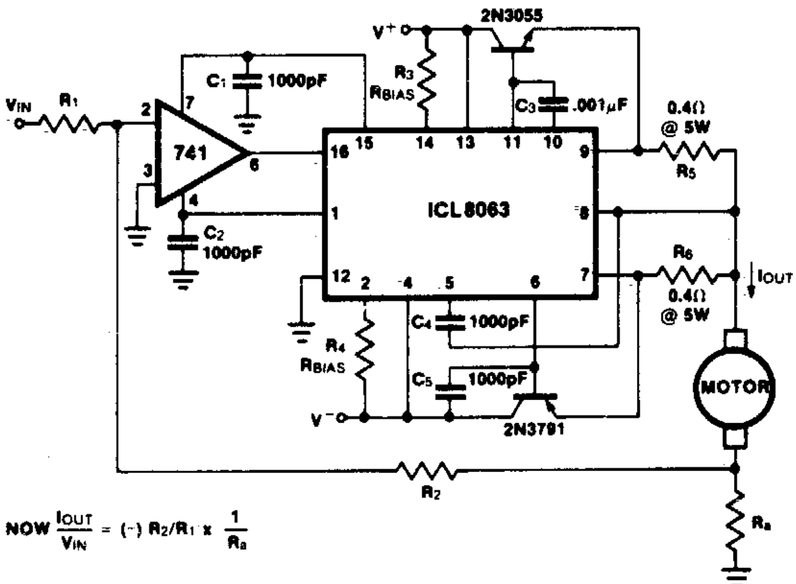


Fig. 57-6

Circuit Notes

Switching action of the 2N4990 allows smaller capacitors to be used while achieving reliable thyristor triggering.

CONSTANT CURRENT MOTOR DRIVE CIRCUIT



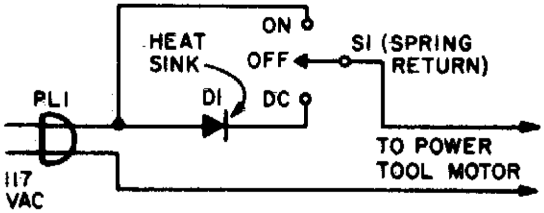
NOW $\frac{I_{OUT}}{V_{IN}} = (-)\ R_2/R_1 \times \frac{1}{R_a}$

Fig. 57-7

Circuit Notes

This minimum device circuit can be used to drive dc motors where there is some likelihood of stalling or lock up; if the motor locks, the current drive remains constant and the system does not destroy itself.

AC MOTOR BRAKE



PARTS LIST FOR
AC MOTOR POWER BRAKE

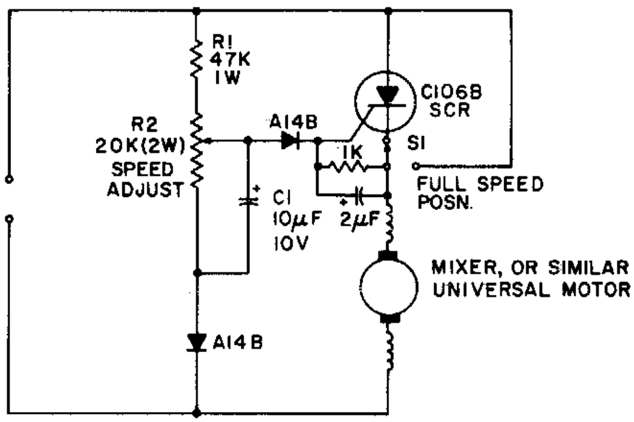
PL1—AC plug
D1—Silicon rectifier, 200 PIV, 20 A.
S1—Spdt switch. Center off, one side spring return
Misc.—Metal cabinet

Circuit Notes

A shot of direct current will instantly stop any ac power tool motor. Switch S1 is a center-off, one side spring return. With S1 on, ac will be fed to the motor and the motor will run. To brake the motor, simply press S1 down and a quick shot of dc will instantly stop it. The switch returns to the center off position when released. This Power Brake can only be used with ac motors; it will not brake universal (ac-dc) motors. A heat sink must be provided for the diode.

Fig. 57-8

UNIVERSAL-MOTOR SPEED CONTROL WITH LOAD-DEPENDENT FEEDBACK (FOR MIXER, SEWING MACHINE, ETC.)



NOTE: RESISTORS 1/2 WATT EXCEPT AS NOTED

Fig. 57-9

Circuit Notes

Simple half-wave motor speed control is effective for use with small universal (ac/dc) motors. Maximum current capability 2.0 amps RMS. Because speed-dependent feedback is provided, the control gives excellent torque characteristics to the motor, even at low rotational speeds. Normal operation at maximum speed can be achieved by closing switch S1, thus bypassing the SCR.

DC MOTOR SPEED/DIRECTION CONTROL CIRCUIT

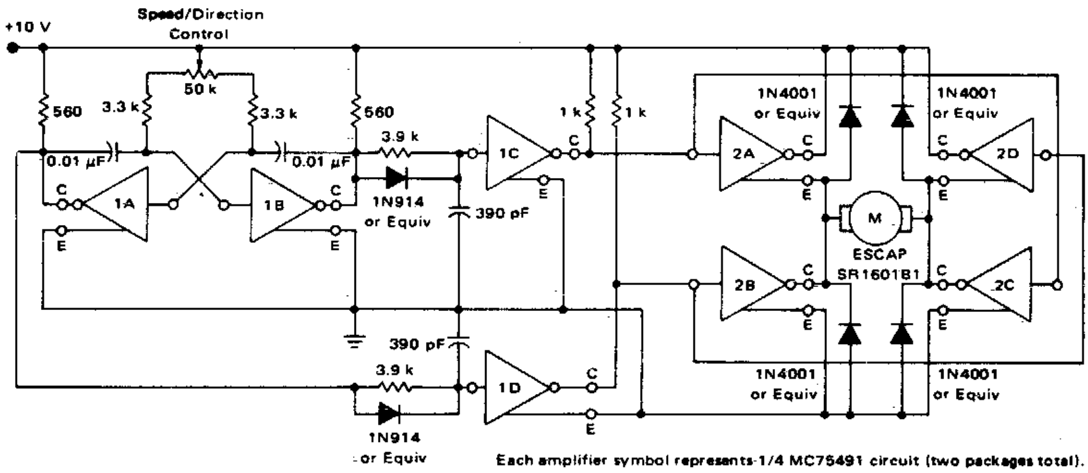
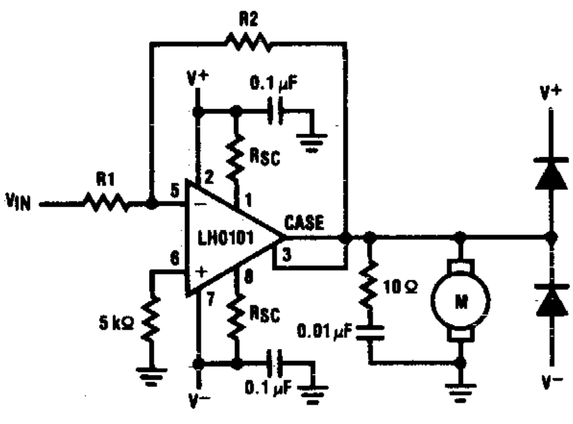


Fig. 57-10

SERVO MOTOR AMPLIFIER

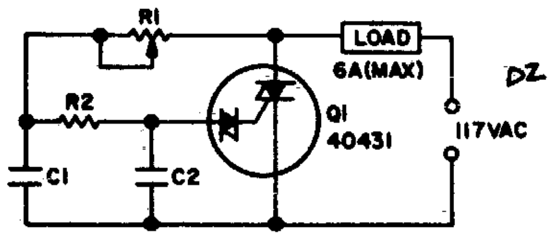


Circuit Notes

Motor driver amplifier will deliver the rated current into the motor. Care should be taken to keep power dissipation within the permitted level. This precision speed regulation circuit employs rate feedback for constant motor current at a given input voltage.

Fig. 57-11

MOTOR SPEED CONTROL



C1, C2—0.1-uF, 200-VDC capacitor
Q1—RCA 40431 Triac-Diac
R1—100,000-ohm linear taper potentiometer
R2—10,000-ohm, 1-watt resistor

Fig. 57-12

Circuit Notes

Universal motors and shaded-pole induction motors can be easily controlled with a full-wave Triac speed controller. Q1 combines both the triac and diac trigger diodes in the same case. The motor used for the load must be limited to 6 amperes maximum. Triac Q1 must be provided with a heat sink. With the component values shown, the Triac controls motor speed from full off to full on.

MODEL TRAIN SPEED CONTROL

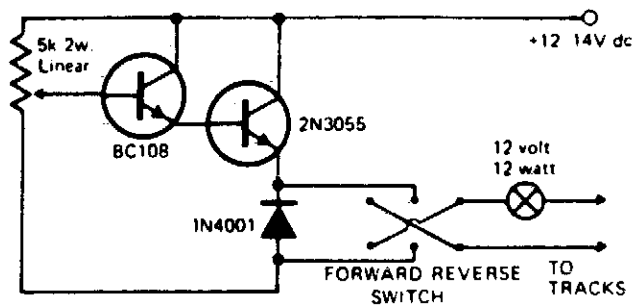


Fig. 57-13

Circuit Notes

Virtually any NPN small signal transistor may be used in place of the BC 108 shown. Likewise any suitable NPN power transistor can be used in place of the 2N3055. The output transistor must be mounted on a suitable heatsink. Short circuit protection may be provided by wiring a 12 volt 12 watt bulb in series with the output. This will glow in event of a short circuit and thus effectively current-limit the output, it also acts as a visual short-circuit alarm.

INDUCTION-MOTOR CONTROL

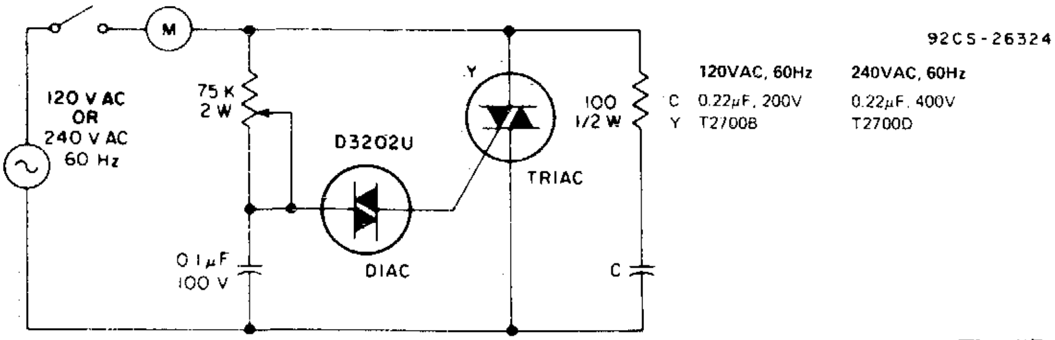


Fig. 57-14

Circuit Notes

This single time-constant circuit can be used as proportional speed control for induction motors such as shaded pole or permanent split-capacitor motors when the load is fixed. The circuit is best suited to applications which require speed control in the medium to full-power range.

DC MOTOR SPEED CONTROL

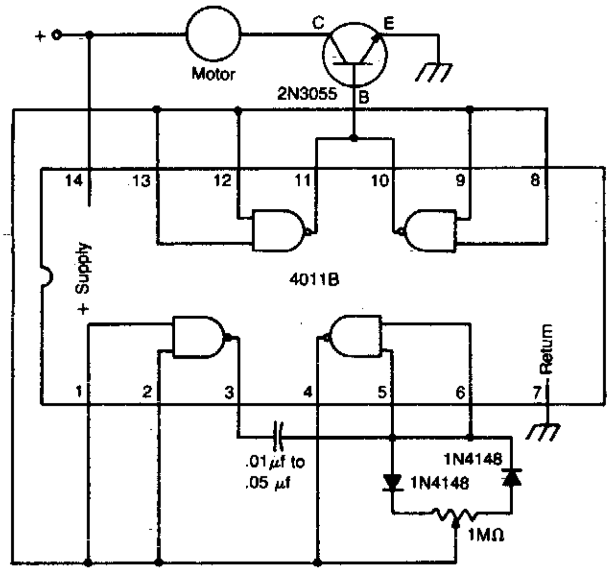


Fig. 57-15

Circuit Notes

The circuit uses a 4011 CMOS NAND gate, a pair of diodes and an NPN power transistor to provide a variable duty-cycle dc source. Adjusting the speed control varies the average voltage applied to the motor. The peak voltage, however, is not changed. This pulse power is effective at very low speeds, constantly kicking the motor along. At higher speeds, the motor behaves in a nearly normal manner.

UNIVERSAL MOTOR CONTROL WITH BUILT-IN SELF TIMER

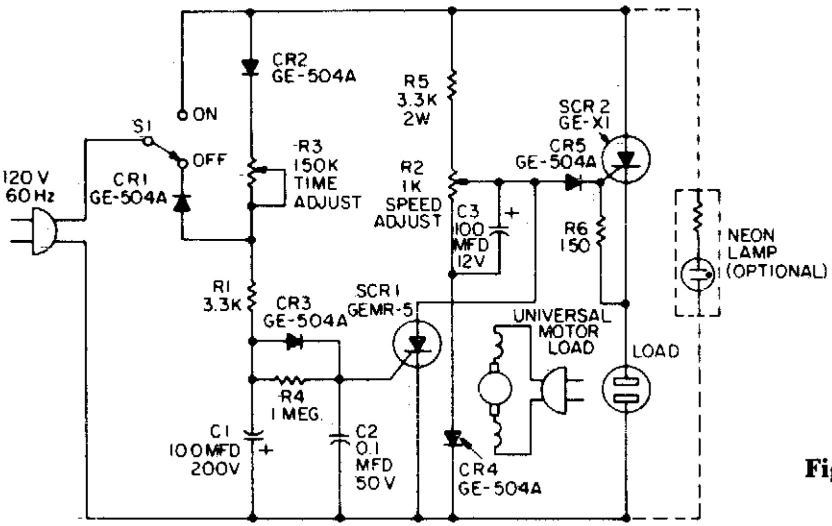


Fig. 57-16

Circuit Notes

When the time delay expires, SCR1 conducts and removes the gate signal from SCR2, which stops the motor. Both the time delay and motor speed are adjustable by potentiometers R2 and R3. If heavier motor loads are anticipated, use the larger C30B SCR in place of the GE-X1 for SCR2. Also, the capacitance of C1 can be increased to lengthen the time delay, if desired.

SPEED CONTROL FOR MODEL TRAINS OR CARS

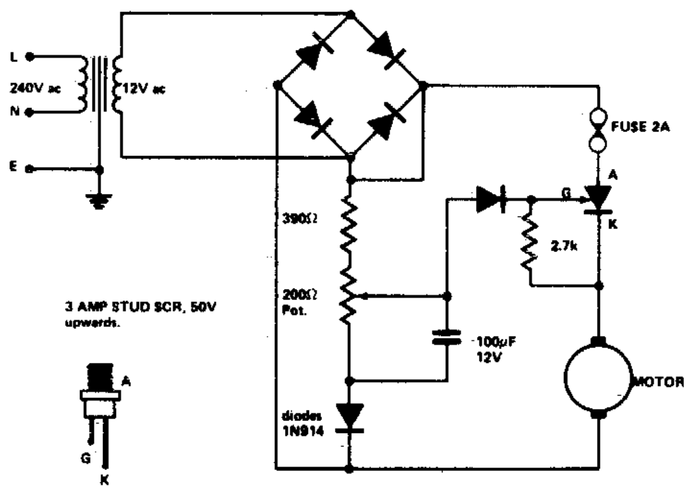


Fig. 57-17

Circuit Notes

Low voltage speed control gives very good starting torque and excellent speed regulation. A reversing switch may be incorporated in the leads to the motor.

DIRECTION AND SPEED CONTROL FOR SHUNT-WOUND MOTORS

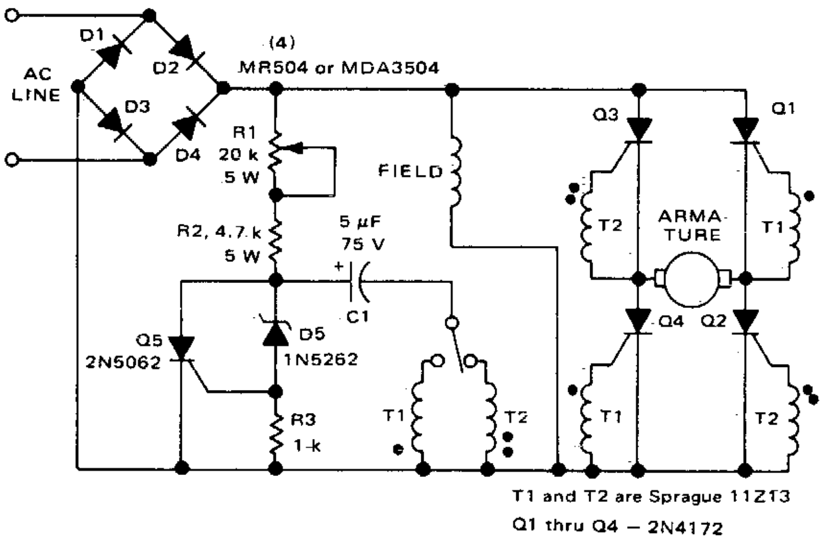


Fig. 57-18

Circuit Notes

This circuit operates like the one shown in Fig. 57-4. The only differences are that the field is placed across the rectified supply and the armature is placed in the SCR bridge. Thus the field current is unidirectional but armature current is reversible; consequently the motor's direction of rotation is reversible. Potentiometer R1 controls the speed.

TWO-PHASE MOTOR DRIVE

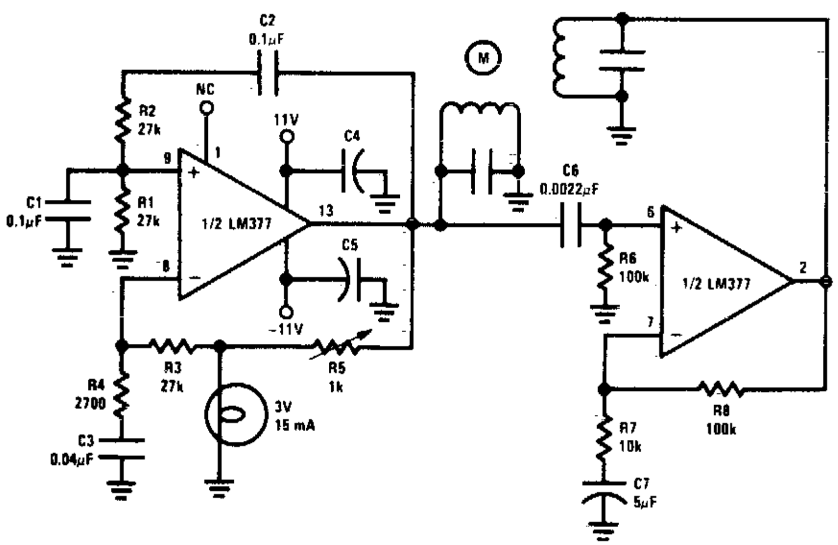


Fig. 57-19

DC SERVO AMPLIFIER

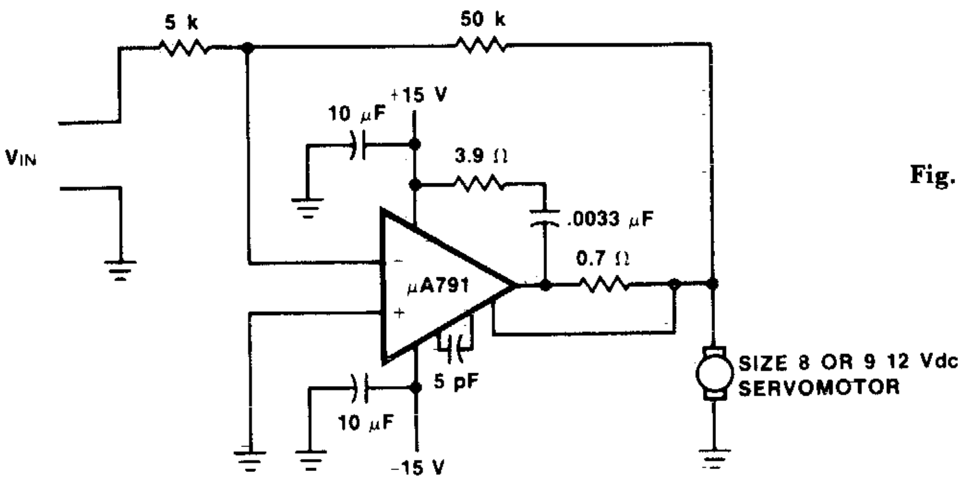
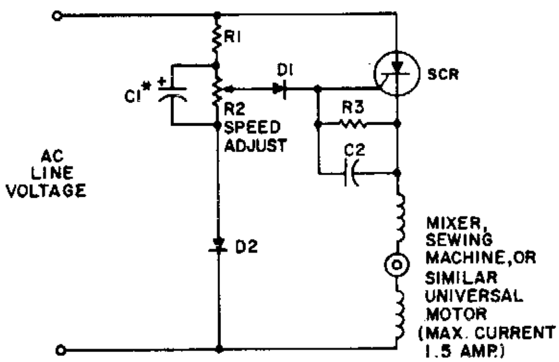


Fig. 57-20

UNIVERSAL MOTOR SPEED CONTROL



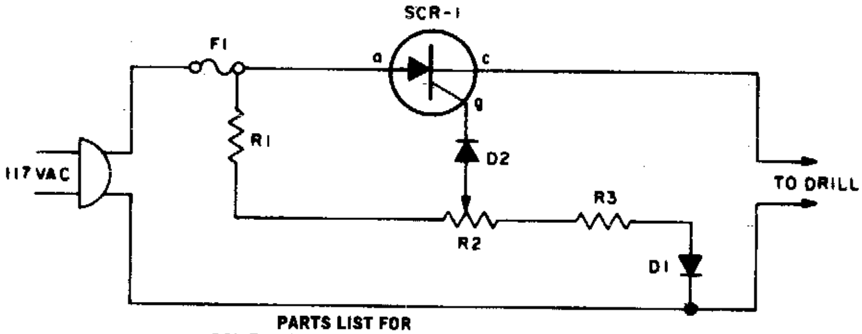
Line Voltage	120V	240V
R_1	47K	100K
R_2	10K	20K
R_3	1K	1K
C_1	1μF, 50V	1μF, 100V
C_2	0.1μF, 50V	0.1μF, 50V
D_1	1N5059	1N5060
D_2	1N5059	1N5060
SCR	C106B1	C106D1

Fig. 57-21

Circuit Notes

The resistor capacitor network R1-R2-C1 provides a ramp-type reference voltage superimposed on top of a dc voltage adjustable with the speed-setting potentiometer R2. This reference voltage appearing at the wiper of R2 is balanced against the residual counter emf of the motor through the SCR gate. As the motor slows down due to heavy loading, its counter emf falls, and the reference ramp triggers the SCR earlier in the ac cycle. More voltage is thereby applied to the motor causing it to pick up speed again. Performance with the C106 SCR is particularly good because the low trigger current requirements of this device allow use of a flat top reference voltage, which provides good feedback gain and close speed regulation.

POWER TOOL TORQUE CONTROL



PARTS LIST FOR
POWER TOOL TORQUE CONTROL

D1, D2—1A, 400 PIV silicon rectifier (Calectro K4-557 or equiv.)
F1—3-A "Slo-blo" fuse
R1—2500-ohm, 5-watt resistor
R2—250-ohm, 4-watt potentiometer
R3—33-ohm, ½-watt resistor
SCR1—8-A, 400-PIV silicon controlled rectifier (HEP R1222)

Fig. 57-22

Circuit Notes

As the speed of an electric drill is decreased by loading, its torque also drops. A compensating speed control like this one puts the oomph back into the motor. When the drill slows down, a back voltage developed across the motor—in series with the SCR cathode and gate—decreases. The SCR gate voltage therefore increases relatively as the back voltage is reduced. The extra gate voltage causes the SCR to conduct over a larger angle and more current is driven into the drill, even as speed falls under load. The SCR should be mounted in ¼-in. thick block of aluminum or copper at least 1-in. square. If the circuit is used for extended periods use a 2 inch square piece.

AC SERVO AMPLIFIER—BRIDGE TYPE

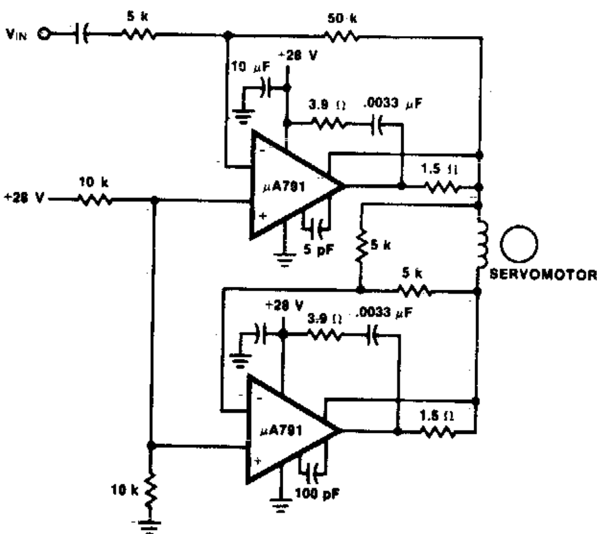


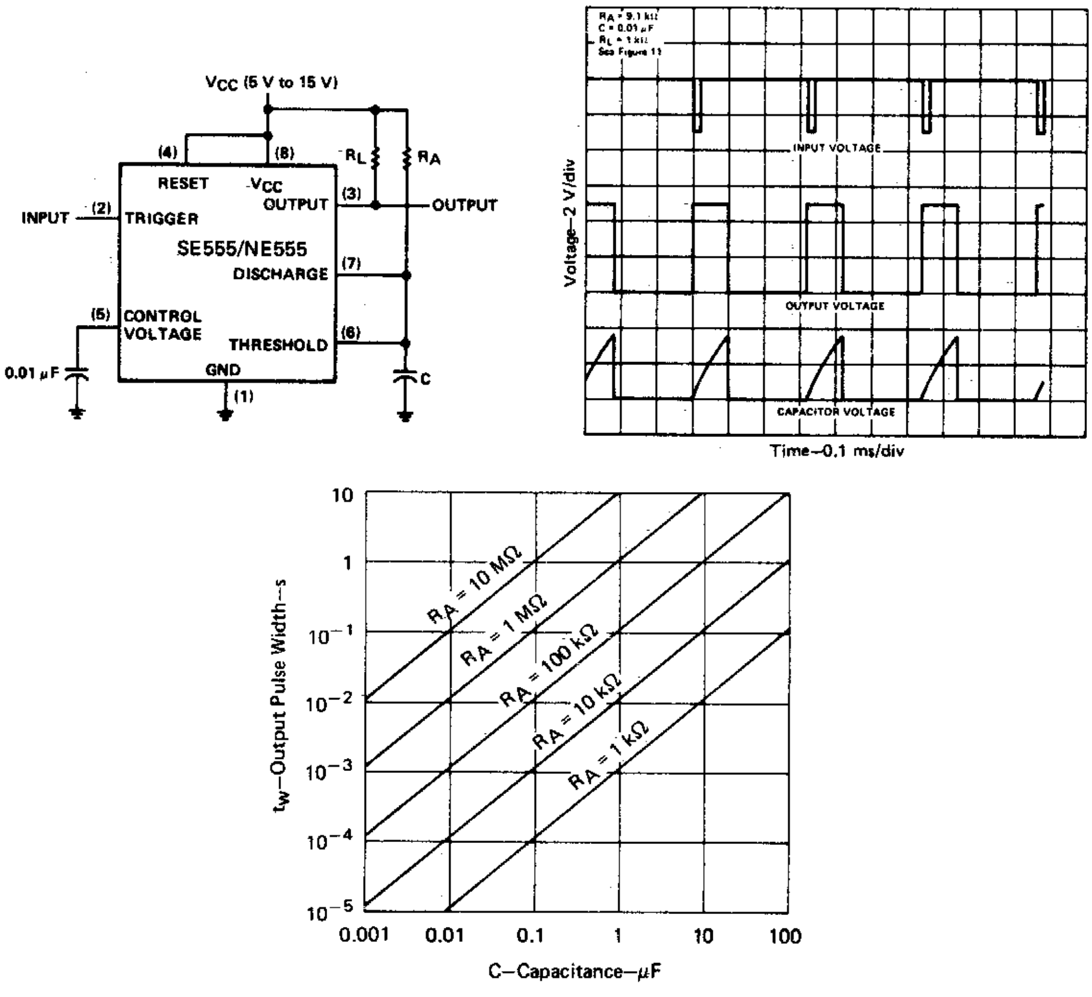
Fig. 57-23

58

Multivibrators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

MONOSTABLE CIRCUIT



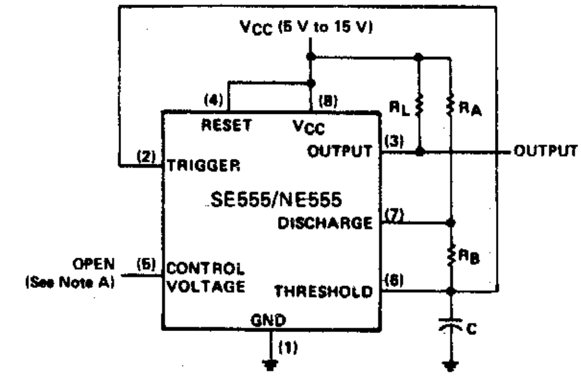
OUTPUT PULSE WIDTH vs CAPACITANCE

Fig. 58-1

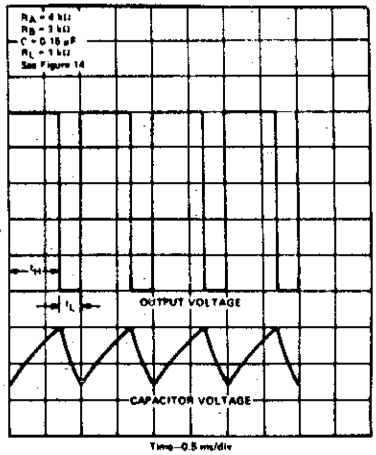
Circuit Notes

If the output is low, application of a negative-going pulse to the trigger input sets the flip-flop (Q goes low), drives the output high, and turns off 1. Capacitor C is then charged through RA until the voltage across the capacitor reaches the threshold voltage of the threshold input. If the trigger input has returned to a high level, the output of the threshold comparator will reset the flip-flop (Q goes high), drive the output low, and discharge C through Q1. Monostable operations is initiated when the trigger input voltage falls below the trigger threshold. Once initiated, the sequence will complete only if the trigger input is high at the end of the timing interval.

ASTABLE MULTIVIBRATOR



NOTE A: Decoupling the control voltage input (pin 5) to ground with a capacitor may improve operation. This should be evaluated for individual applications.



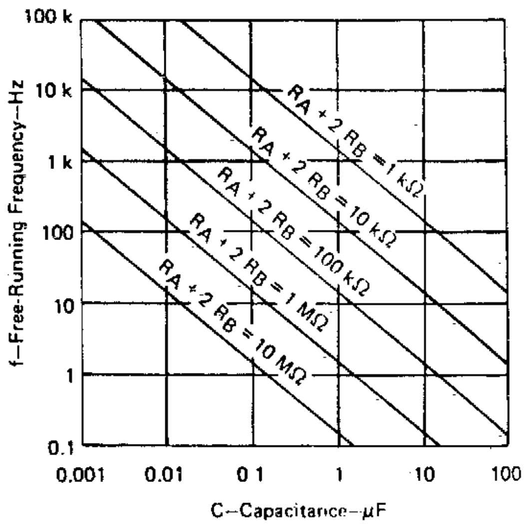


Fig. 58-2

Circuit Notes

The capacitor C will charge through R_A and R_B, and then discharge through R_B only. The duty cycle may be controlled by the values of R_A and R_B.

ASTABLE OSCILLATOR

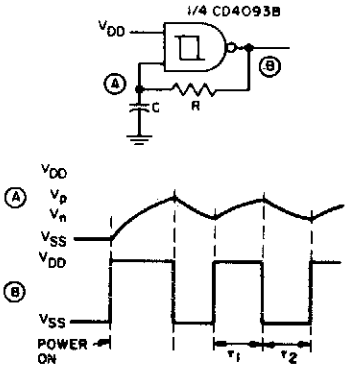
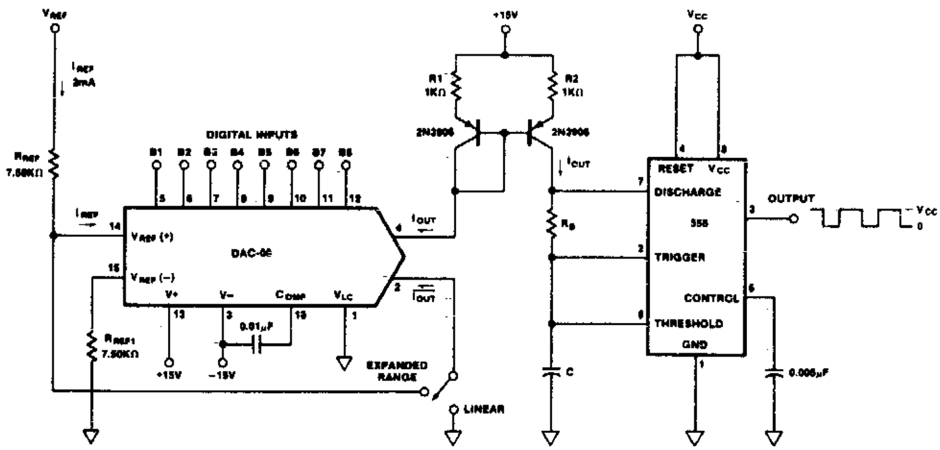


Fig. 58-3

Circuit Notes

Before power is applied, the input and output are at ground potential and capacitor C is discharged. On power-on, the output goes high (V_{DD}) and C charges through R until V is reached; the output then goes low (V_{SS}). C is now discharged through R until V_n is reached. The output then goes high and charges C towards V_p through R. Thus input A alternately swings between V_p and V_n as the output goes high and low. This circuit is self-starting at power-on.

DIGITALLY CONTROLLED ASTABLE MULTIVIBRATOR

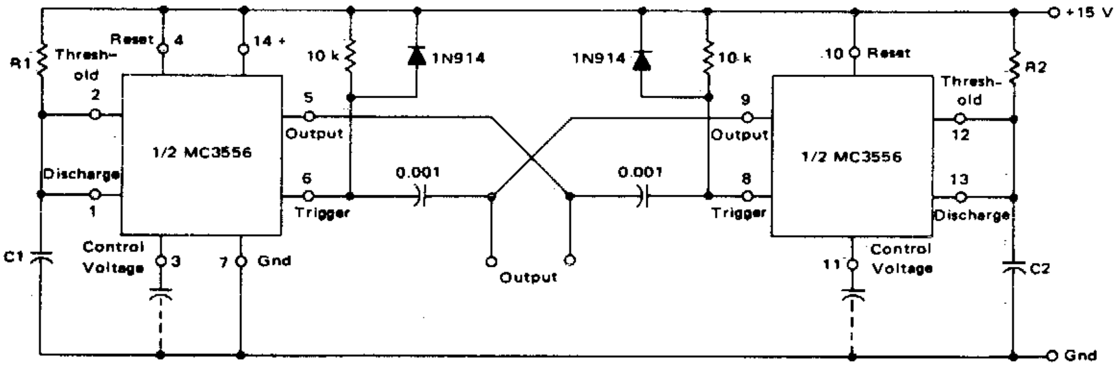


$$\text{FREQUENCY, } f = \frac{1}{\frac{1}{3}\,\frac{R_{REF}C}{\{D\}}\,\frac{V_{CC}}{V_{REF}} + 0.695\,R_BC} \quad \text{FOR LINEAR MODE}$$

$$\text{FREQUENCY, } f = \frac{1}{\frac{1}{3}\,R_{REF}C\,\frac{V_{CC}}{V_{REF}}\left[\frac{2-\{D\}}{\{D\}}\right] + 0.695\,R_BC} \quad \text{FOR EXPANDED MODE}$$

Fig. 58-4

DUAL ASTABLE MULTIVIBRATOR



$f = \frac{0.91}{(R1+R2)\,C}$ for C1 = C2　　Duty Cycle $\frac{R2}{R1+R2}$

Fig. 58-5

Circuit Notes

This dual astable multivibrator provides versatility not available with single timer circuits. The duty cycle can be adjusted from 5% to 95%. The two outputs provide two phase clock signals often required in digital systems. It can also be inhibited by use of either reset terminal.

UJT MONOSTABLE

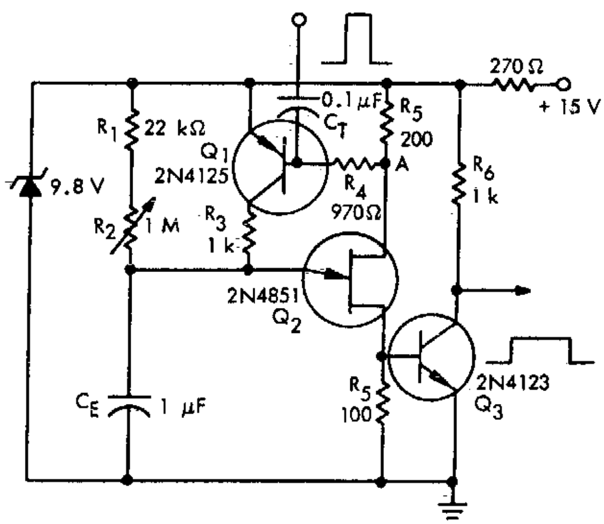


Fig. 58-6

MONOSTABLE MULTIVIBRATOR WITH INPUT LOCK-OUT

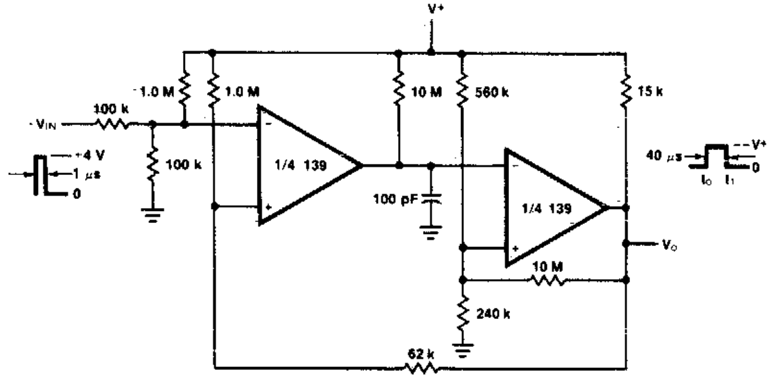


Fig. 58-7

TTL MONOSTABLE

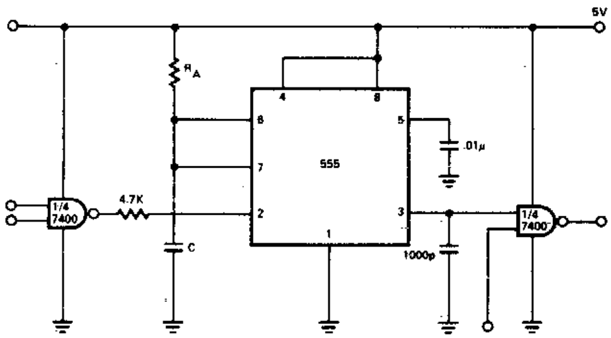


Fig. 58-8

MONOSTABLE CIRCUIT

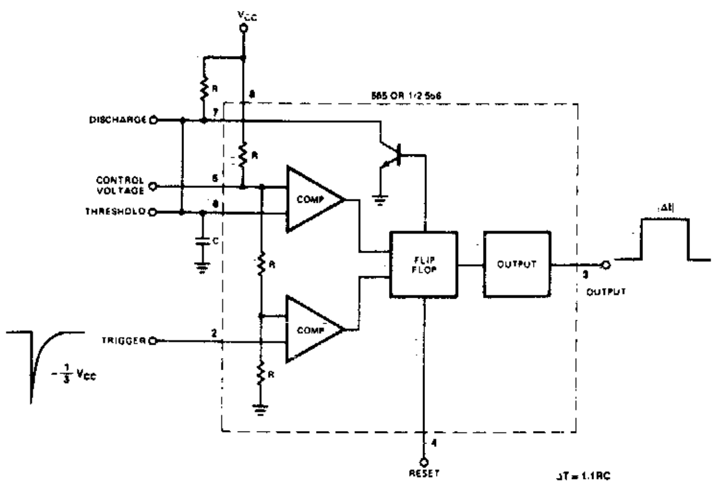


Fig. 58-9

ONE-SHOT MULTIVIBRATOR

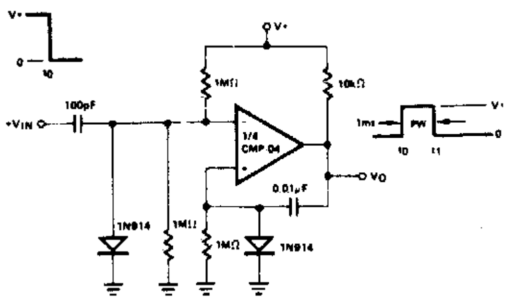


Fig. 58-10

BISTABLE MULTIVIBRATOR

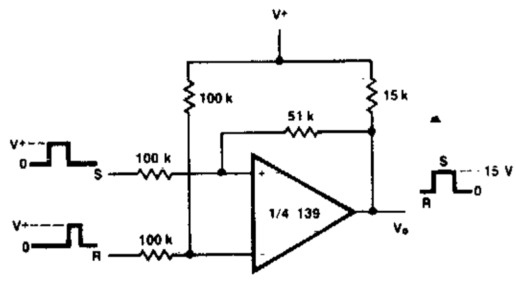


Fig. 58-12

MONOSTABLE MULTIVIBRATOR

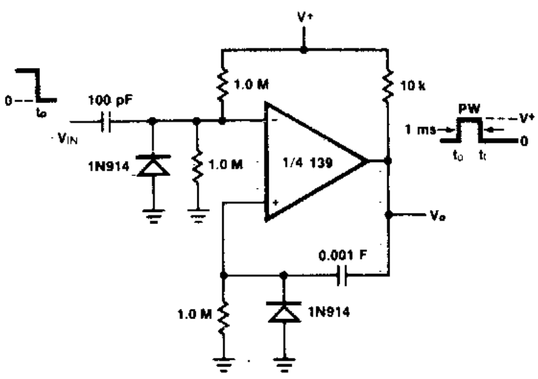
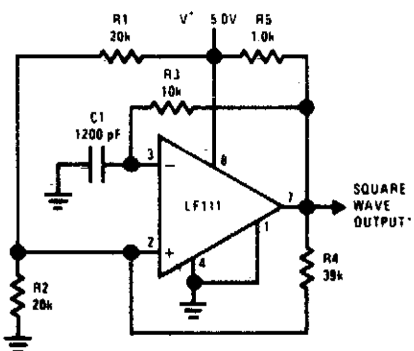


Fig. 58-11

100 kHz FREE-RUNNING MULTIVIBRATOR



*TTL or DTL fanout of two.

Fig. 58-13

59

Noise Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AUDIO NOISE GENERATOR

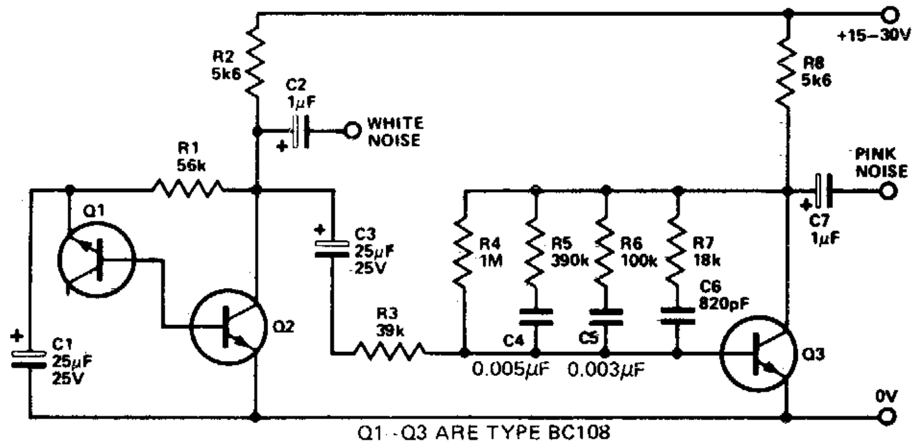


Fig. 59-1

Circuit Notes

This simple circuit generates both white and pin noise. Transistor Q1 is used as a zener diode. The normal base-emitter junction is reverse-biased and goes into zener breakdown at about 7 to 8 volts. The zener noise current from Q1 flows into the base of Q2 such that an output of about 150 millivolts of white noise is available. To convert the white noise to pink, a filter is required which provides a 3 dB cut per octave as the frequency increases. Since such a filter attenuates the noise considerably an amplifier is used to restore the output level. Transistor Q3 is this amplifier and the pink noise filter is connected as a feedback network. between collector and base in order to obtain the required characteristic by controlling the gain-versus-frequency of the transistor. The output of transistor Q3 is thus the pink noise required and is fed to the relevant output socket.

PINK NOISE GENERATOR

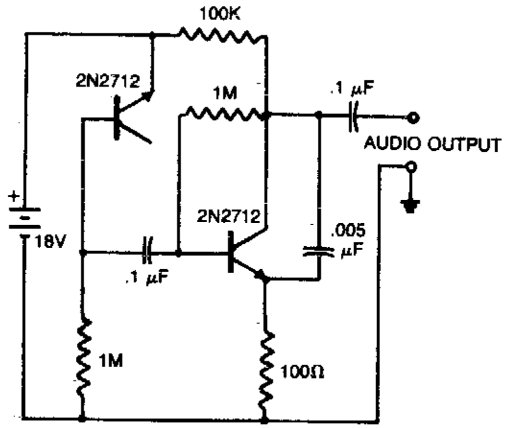


Fig. 59-2

Circuit Notes

A reverse-biased pn junction of a 2N2712 transistor is used as a noise generator. The second 2N2712 is an audio amplifier. The 0.005 μF capacitor across the amplifier output removes some high-frequency components to simulate pink noise more closely. The audio output may be connected to high-impedance earphones or to a driver amplifier for speaker listening.

NOISE GENERATOR

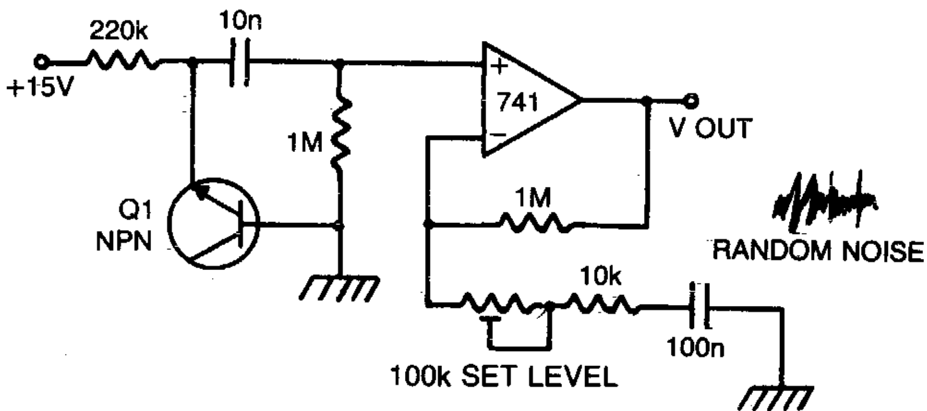


Fig. 59-3

Circuit Notes

The zener breakdown of a transistor junction is used as a noise generator. The breakdown mechanism is random and this voltage has a high source impedance. By using the op amp as a high input impedance, high ac gain amplifier, a low impedance, large signal noise source is obtained. The 100K potentiometer is used to set the noise level by varying the gain from 40 to 20 dB.

WIDEBAND NOISE GENERATOR

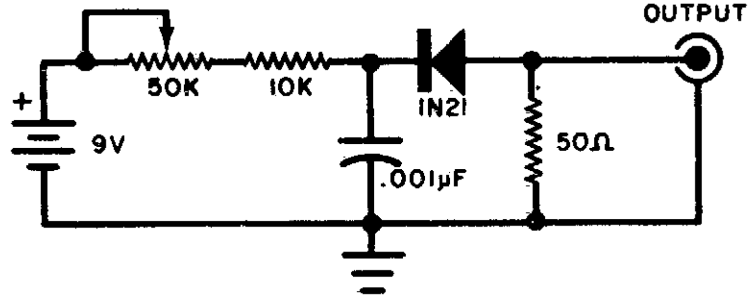


Fig. 59-4

Circuit Notes

This circuit will produce wideband rf noise. It uses a reverse-biased diode and has a low-impedance output. Can be used to align receivers for optimum performance.

NOISE GENERATOR CIRCUIT

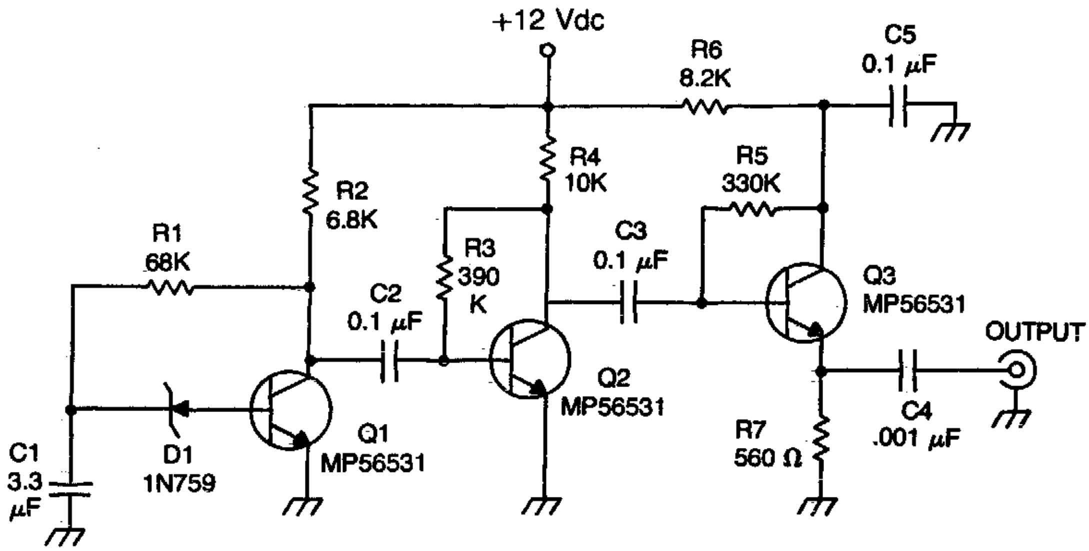


Fig. 59-5

Circuit Notes

The zener diode is an avalanche rectifier in the reverse bias mode connected to the input circuit of a wideband rf amplifier. The noise is amplified and applied to the cascade wideband amplifier, transistors Q2 and Q3.

60

Oscilloscope Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

OSCILLOSCOPE CONVERTER PROVIDES FOUR-CHANNEL DISPLAYS

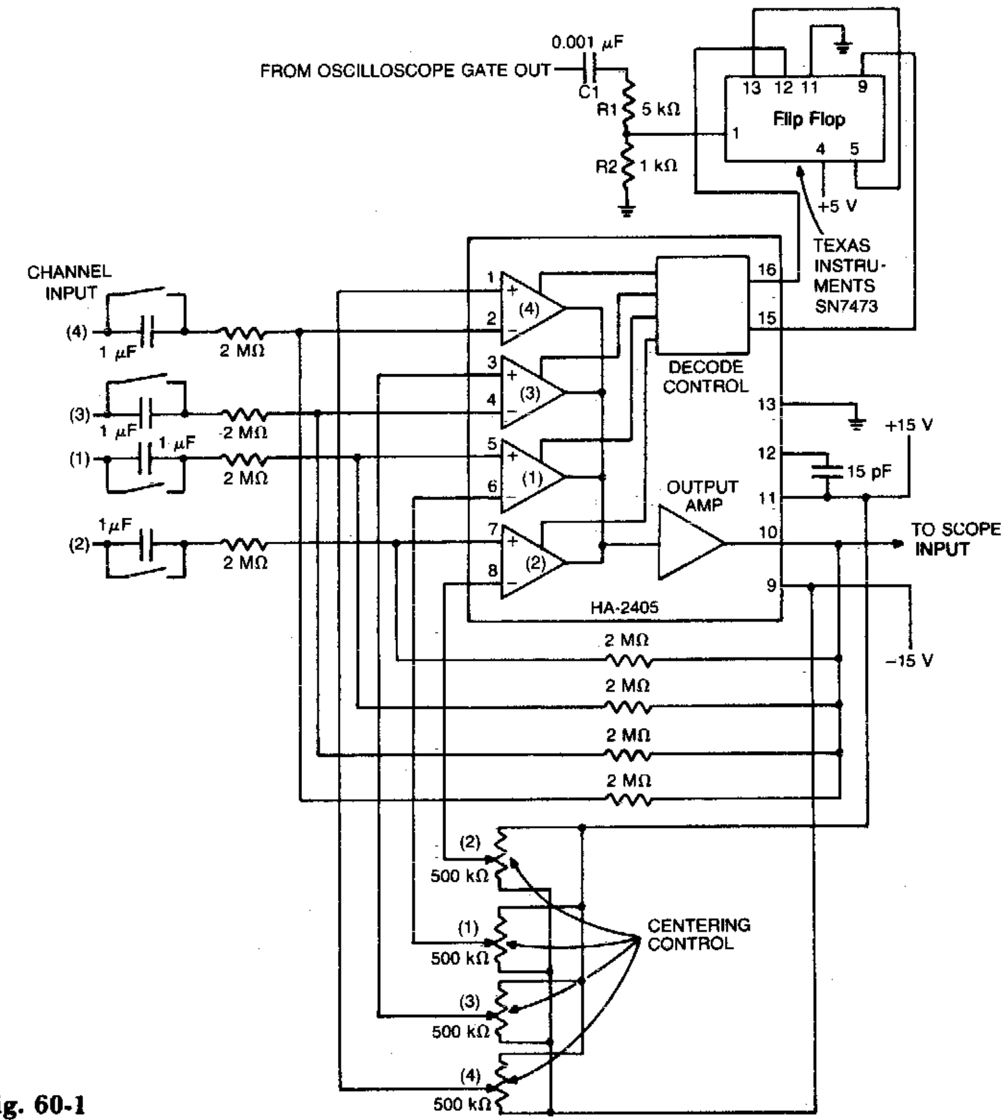


Fig. 60-1

Circuit Notes

The monolithic quad operational amplifier provides an inexpensive way to increase display capability of a standard oscilloscope. Binary inputs drive the IC op amp; a dual flip-flop divides the scope's gate output to obtain channel selection signals. All channels have centering controls for nulling offset voltage. A negative-going scope gate signal selects the next channel after each trace. The circuit operates out to 5 MHz.

ADD-ON TRIGGERED SWEEP

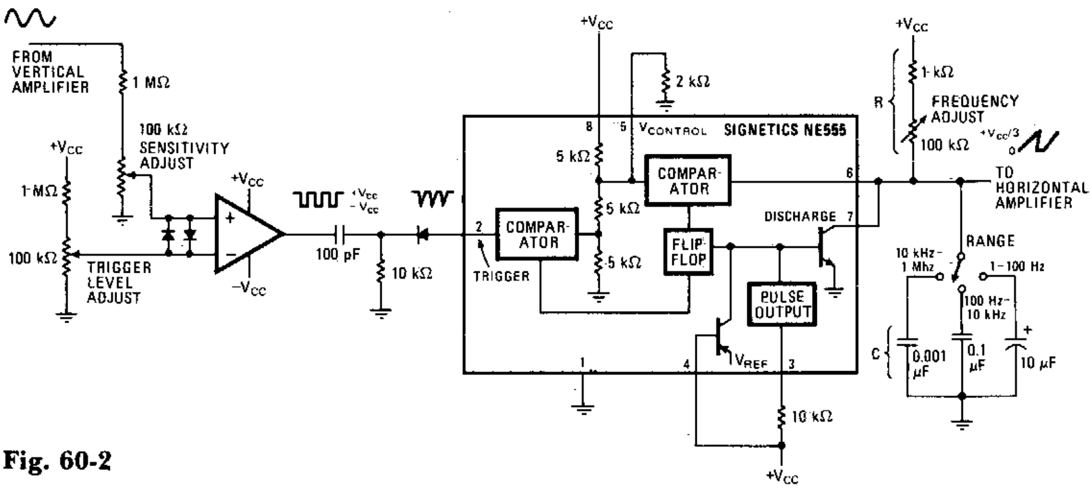


Fig. 60-2

Circuit Notes

The circuit's input op amp triggers the timer, setting its flip-flop and cutting off its discharge transistor so that capacitor C can charge. When the capacitor voltage reaches the timer's control voltage (0.33Vcc), the flip-flop resets and the transistor conducts, discharging the capacitor.

10.7 MHz SWEEP GENERATOR

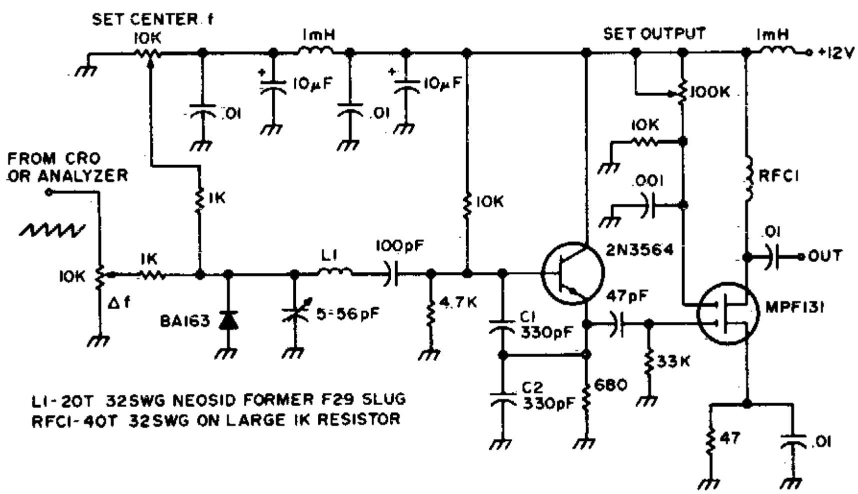


Fig. 60-3

Circuit Notes

This circuit is used to observe the response of an if amp or a filter. It can be used with an oscilloscope or, for more dynamic range, with a spectrum analyzer.

DRAWING CIRCLES ON A SCOPE

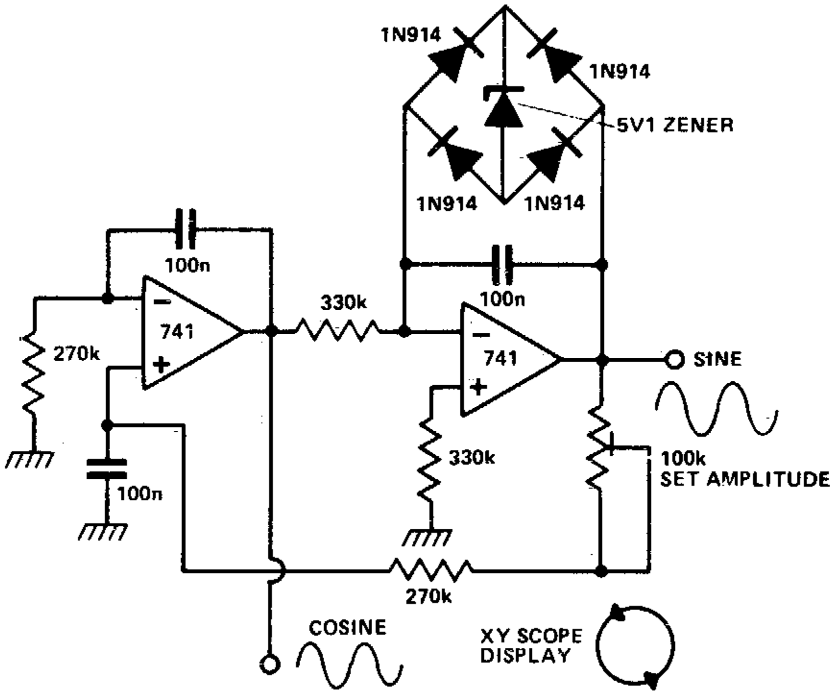


Fig. 60-4

Circuit Notes

The circuit is that of a quadrature sine and cosine oscillator. To generate circular displays, connect the two outputs to the X and Y inputs.

TRANSMITTER-OSCILLOSCOPE COUPLER FOR CB SIGNALS

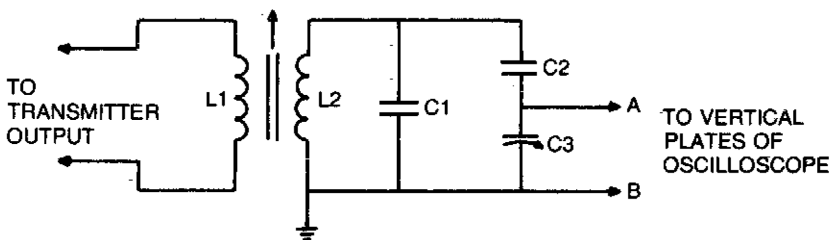


Fig. 60-5

Circuit Notes

To display an rf signal, connect L1 to the transmitter and points A and B to the vertical plates of the oscilloscope. Adjust L1 for minimum SWR and C3 for the desired trace height on the CRT. L2 = 4 turns #18 on ¾" slug tuned rf coil form, L1 = 3 turns #22 adjacent to grounded end of L1, C1, and C2 = 5 pF, C3 = 75 pF trimmer.

OSCILLOSCOPE MONITOR

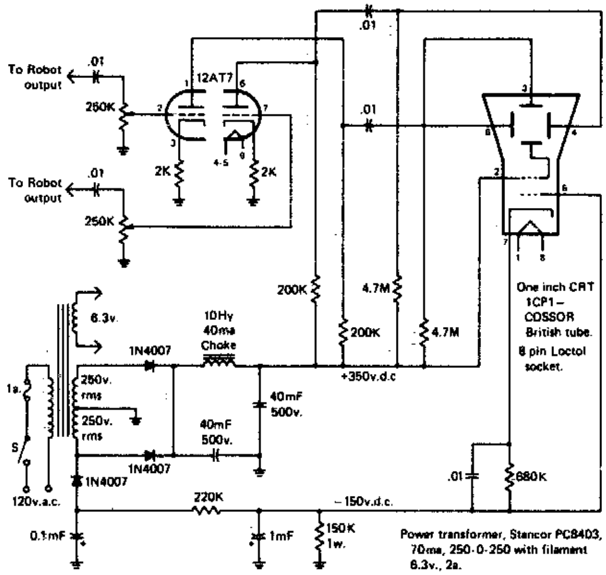


Fig. 60-6

BEAM SPLITTER FOR OSCILLOSCOPE

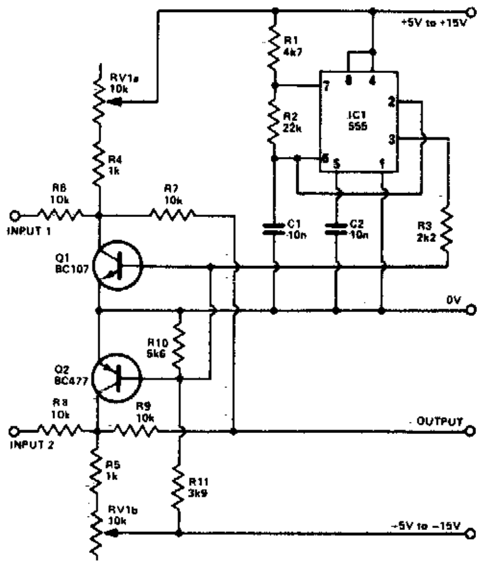


Fig. 60-7

Circuit Notes

The basis of the beam-splitter is a 555 timer connected as an astable multivibrator. Signals at the two inputs are alternately displayed on the oscilloscope with a clear separation between them. The output is controlled by the tandem potentiometer RV1a/b which also varies the amplitude of the traces.

61

Phase Sequence and Phase Shift Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PHASE SEQUENCE INDICATOR

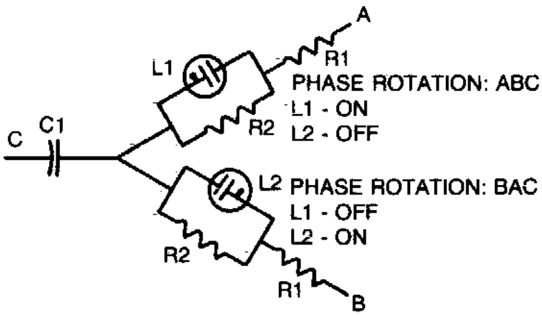


Fig. 61-1

Circuit Notes

Simple, portable phase-sequence indicator determines the proper phase rotation in polyphase circuits. Major components are two neon lamps, two resistors, and a capacitor. In operation, the leg voltages are unbalanced, so that the lamp with the maximum voltage—or proper phase sequence—lights. Table shows typical component values for various circuit frequencies.

SINGLE TRANSISTOR PHASE SHIFTER

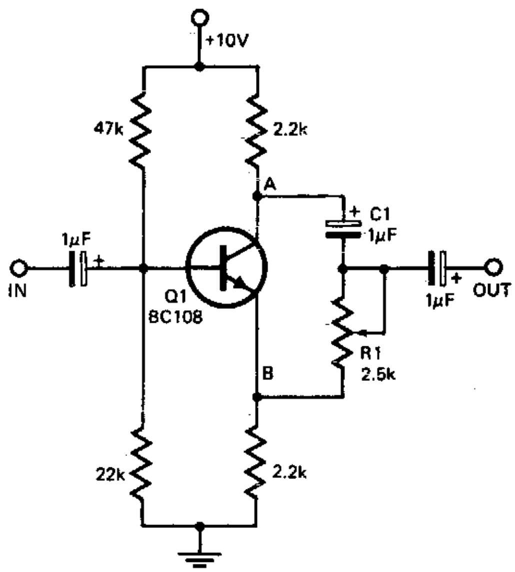
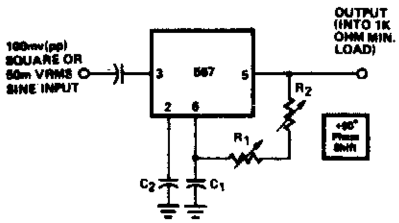


Fig. 61-2

Circuit Notes

This circuit provides a simple means of obtaining phase shifts between zero and 170°. The transistor operates as a phase splitter, the output at point A being 180° out of phase with the input. Point B is in phase with the input phase. Adjusting R1 provides the sum of various proportions of these and hence a continuously variable phase shift is provided. The circuit operates well in the 600 Hz to 4 kHz range.

0° TO 180° PHASE SHIFTER



$R_2 = R_1/5$

Adjust R_1 so that $\phi = 90°$ with control midway

Fig. 61-3

PRECISION PHASE SPLITTER

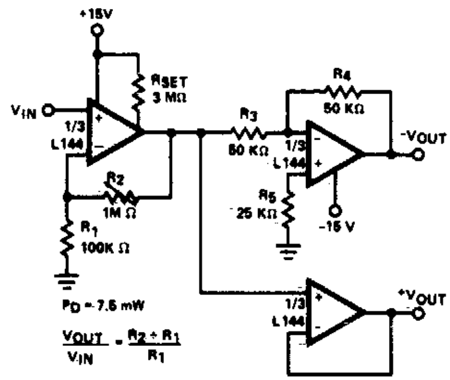
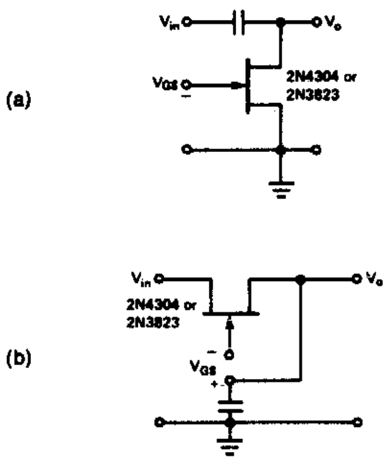


Fig. 61-5

PHASE SHIFT CIRCUITS



(a) Phase advance circuit.

(b) Phase retard circuit.

Fig. 61-4

0° TO 360° PHASE SHIFTER

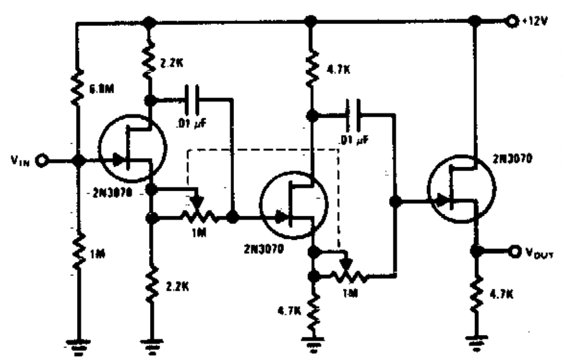


Fig. 61-6

Circuit Notes

Each stage provides 0° to 180° phase shift. By ganging the two stages, 0° to 360° phase shift is achieved. The 2N3070 JFETs do not load the phase shift networks.

62

Photography Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Automatic Contrast Meter
- Darkroom Timer
- Photo Stop Action
- Sound Light-Flash Trigger
- Sound Activated Strobe Trip
- Flash Slave Driver
- Remote Flash Trigger
- Flash Exposure Meter
- Shutter Tester
- Photographic Timer

AUTOMATIC CONTRAST METER

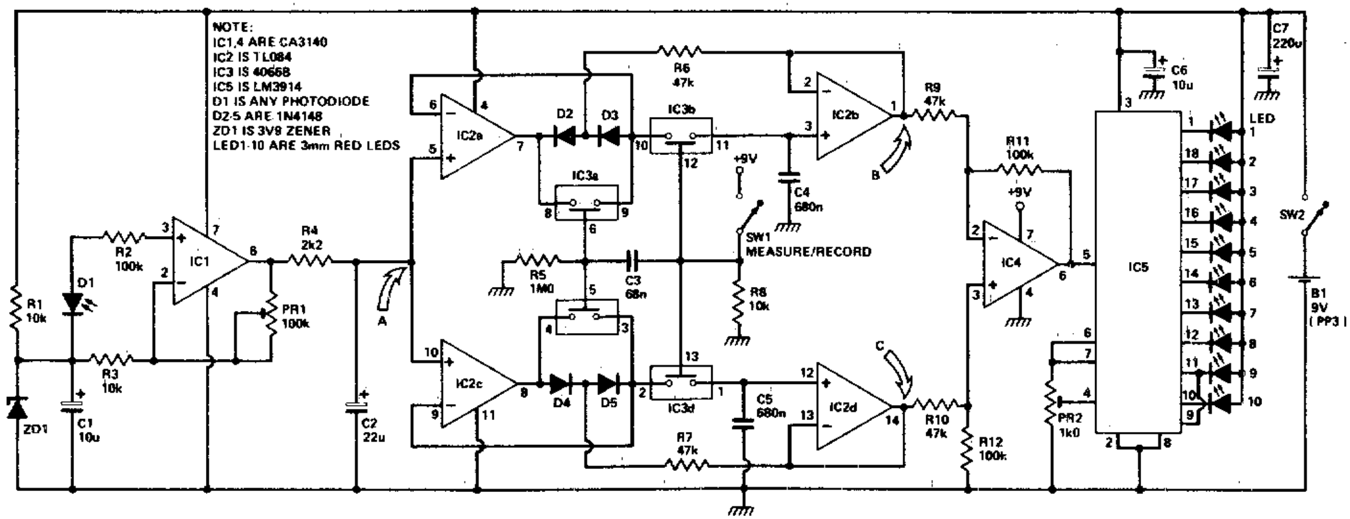


Fig. 62-1

Circuit Notes

The circuit arrangement consists of a photo-amplifier which feeds a voltage derived from varying light levels in an enlarger to a pair of peak detectors. One follows the peak positive voltage and the other the peak negative voltage. The capacitors used for storing the voltage peaks in the followers also form part of sample and hold circuits which are then switched to hold after the measurement. Their outputs represent the maximum and minimum values of light intensity. A differential amplifier then computes the ratio of these values, and the result is displayed on an LED bargraph meter.

DARKROOM TIMER

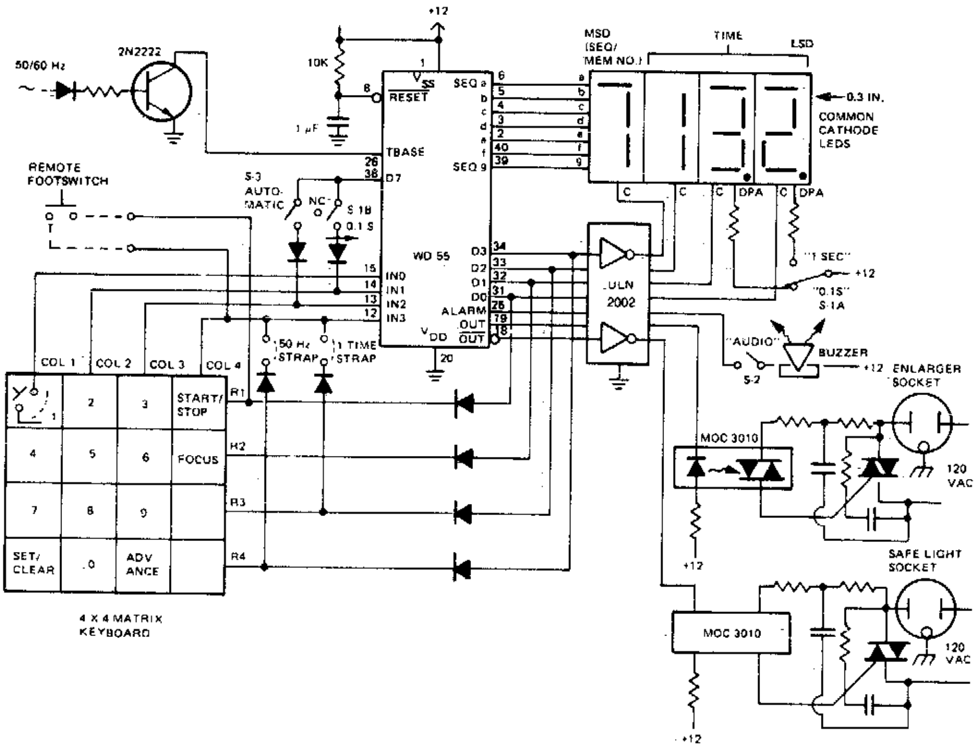


Fig. 62-2

Circuit Notes

The darkroom timer/controller uses few external components: a display, a digit driver, keyboard, and output switching devices. A 4-digit common-cathode LED display is desirable for dark room environments. The time base is provided by shaping up the 50/60 Hz ac line. A DPDT switch (S1) is used to select a resolution of .1 or 1 seconds and to simultaneously move the decimal point. Timer/controller has two switched ac outlets, one for the enlarger and one for the safe light. They are the complements of each other in that the safe light is on when the enlarger is not active and is off when the enlarger is printing. The buzzer is of the self-contained oscillator variety and operates with dc drive.

PHOTO STOP ACTION

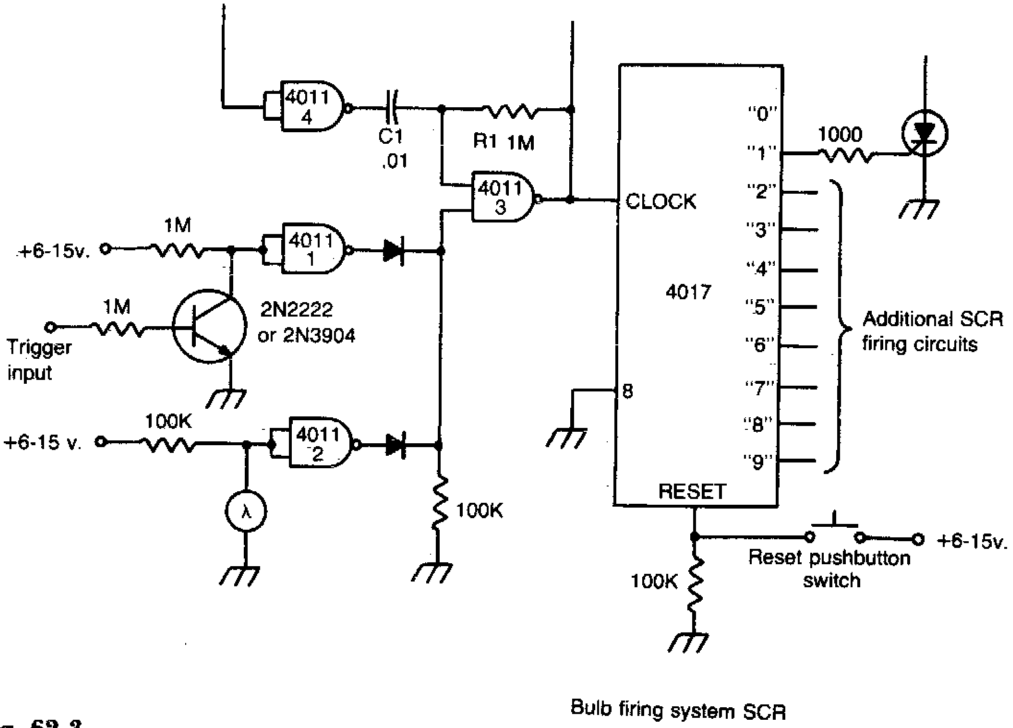


Fig. 62-3

Circuit Notes

This circuit gives multiple "stop-action" photographic effects like showing a bouncing ball in up to nine locations in a single photograph. The circuit will automatically fire the bulbs sequentially with the time between each firing variable. The circuit is functionally complete except for the actual firing system. In many cases, a simple SCR will work, as shown. The firing can be initiated in one of two ways. A trigger pulse can be applied to the trigger input terminal through a capacitor, or can operate the unit as a slave. Light from a camera-mounted flash will activate the circuit through its built-in photocell pickup. The time period between each successive flash is determined by C1 and R1, which is variable. After firing the circuit, it must be reset by momentarily depressing the reset button.

SOUND LIGHT-FLASH TRIGGER

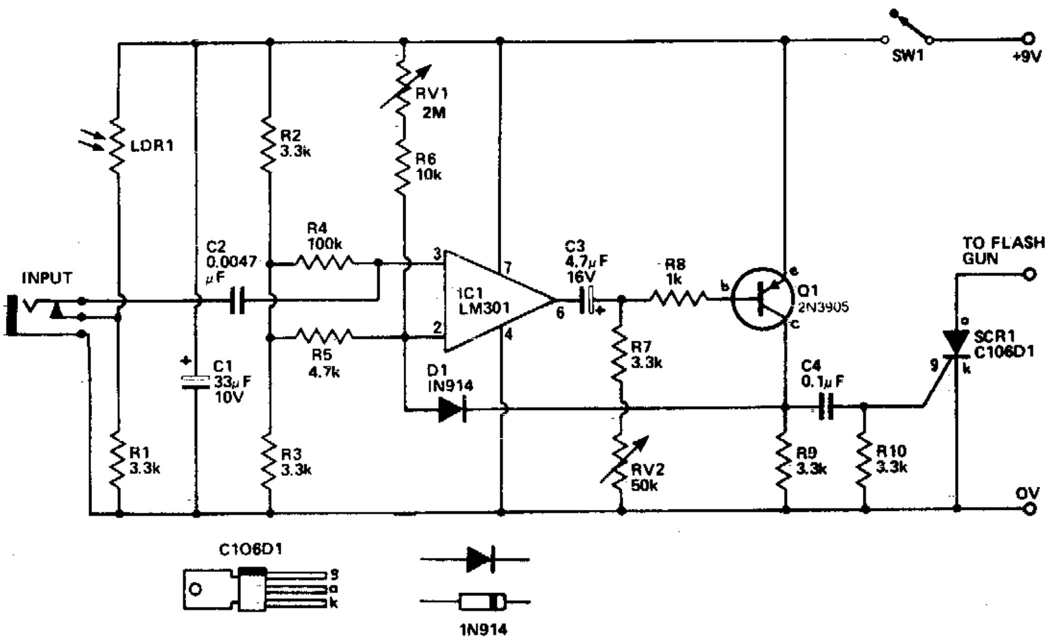
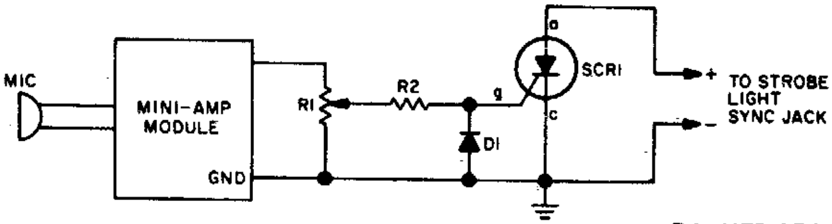


Fig. 62-4

Circuit Notes

Sound input to the microphone triggers the IC monostable circuit which subsequently triggers an SCR, and hence the flash, after a time delay. This delay is adjustable—by varying the monostable on-time—from from 5 milliseconds to 200 milliseconds.

SOUND ACTIVATED STROBE TRIP



D1—HEP-154 silicon rectifier
R1—5000-ohm potentiometer
R2—2700-ohm, ½-watt resistor
SCR1— silicon- controlled rectifier
MIC.—Ceramic microphone

Fig. 62-5

Circuit Notes

Take strobe-flash pictures the instant a pin pricks a balloon, a hammer breaks a lamp bulb or a bullet leaves a gun. Use a transistor amplifier of 1-watt rating or less. (It must have an output transformer.) The amplifier is terminated with a resistor on its highest output impedance, preferably 16 ohms. To test, darken room lights, open camera shutter, and break a lamp bulb with a hammer. The sound of the hammer striking the lamp will trigger the flash, and the picture will have been taken at that instant.

FLASH SLAVE DRIVER

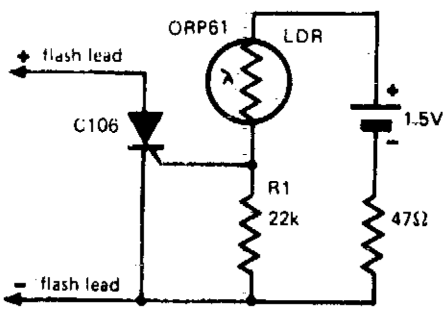


Fig. 62-6

Circuit Notes

In photography, a separate flash, triggered by the light of a master flash light, is often required to provide more light, fill-in shadows etc. The sensitivity of this circuit depends on the proximity of the master flash and the value of R1. Increasing R1 gives increased sensitivity.

REMOTE FLASH TRIGGER

Q1—300-V light-activated silicon-controlled rectifier (LASCR)
R1—47,000-ohm, ½-watt resistor

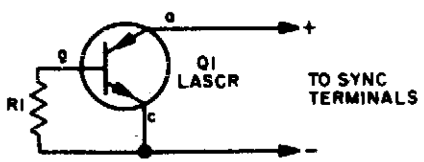


Fig. 62-7

Circuit Notes

Transistor Q1 is a light-activated silicon-controlled rectifier (LASCR). The gate is tripped by light entering a small lens built into the top cap. To operate, provide a 6-in. length of stiff wire for the anode and cathode connections and terminate the wires in a polarized power plug that matches the sync terminals on your electronic flashgun (strobelight). Make certain the anode lead connects to the positive sync terminal. When using the device, bend the connecting wires so the LASCR lens faces the main flash. This will fire the remote unit.

FLASH EXPOSURE METER

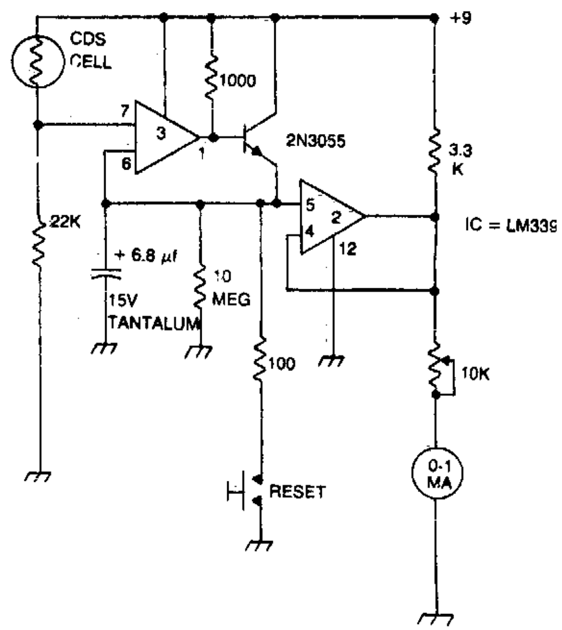


Fig. 62-8

Circuit Notes

Strobe light meter catches the peak of flash intensity and holds it long enough to give a reading. The reset button must be pressed before each measurement.

SHUTTER TESTER

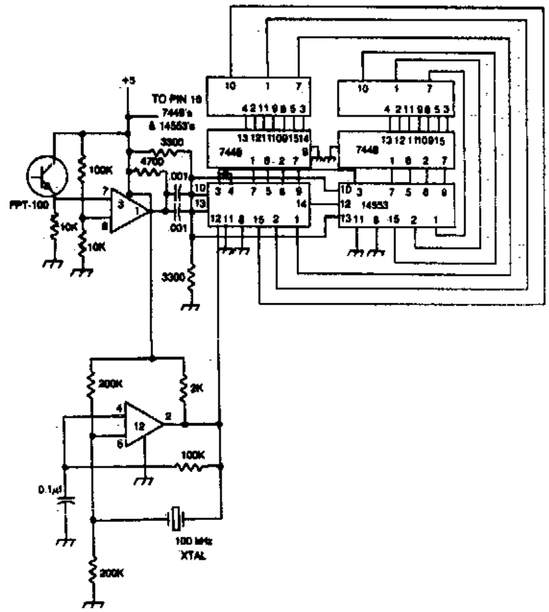


Fig. 62-9

Circuit Notes

Shutter speed tester combines frequency counter, crystal oscillator, and phototransistor-operated gate generator. Oscillator pulses are counted as long as the shutter is open. Reset is automatic at the instant the shutter opens.

PHOTOGRAPHIC TIMER

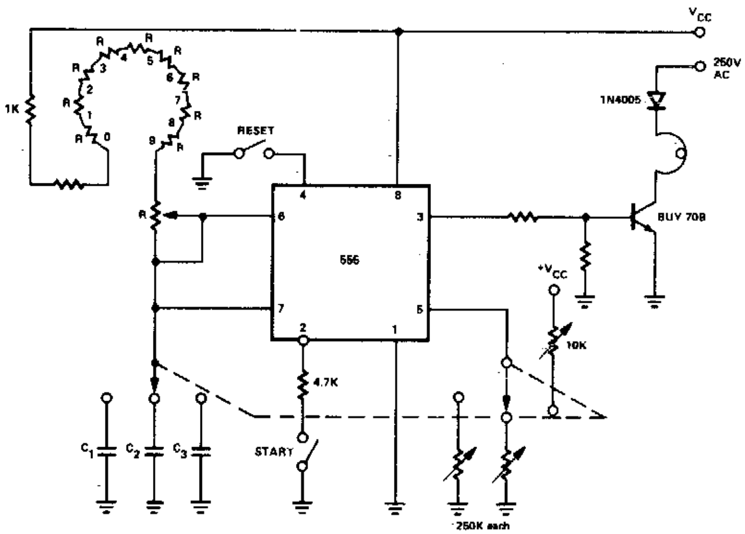


Fig. 62-10

63

Power Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

EXTENDED RANGE VU METER (DOT MODE)

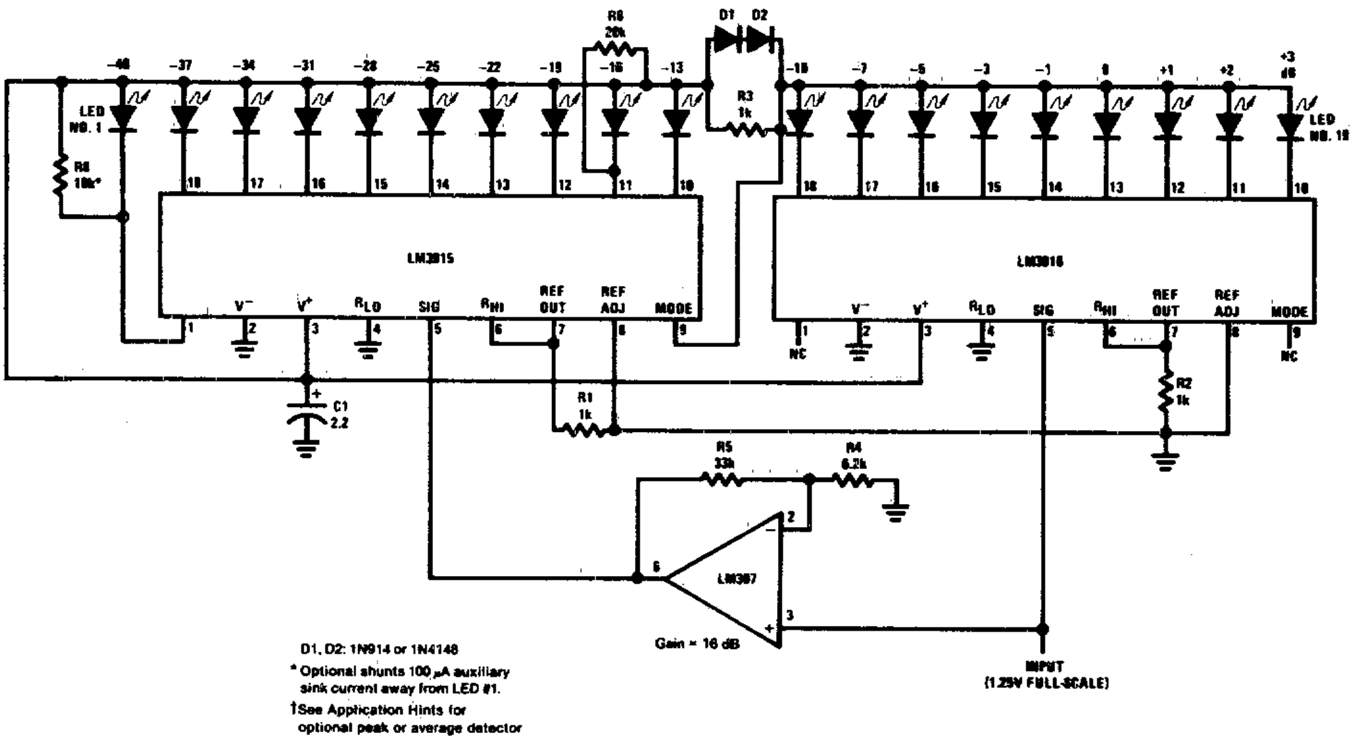
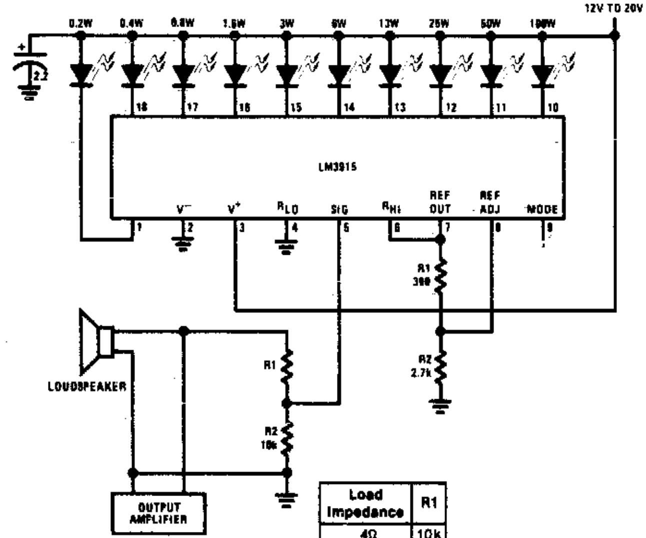


Fig. 63-1

AUDIO POWER METER



Load Impedance	R1
4Ω	10k
8Ω	18k
16Ω	30k

See Application Hints for optional Peak or Average Detector

Fig. 63-2

AUDIO POWER METER

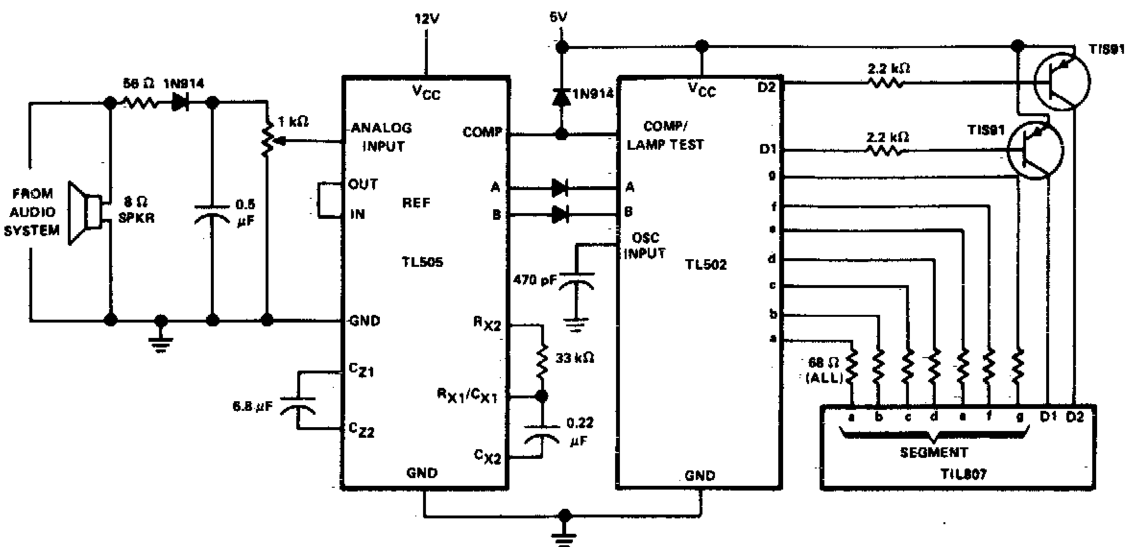


Fig. 63-3

POWER METER (1 kW FULL SCALE)

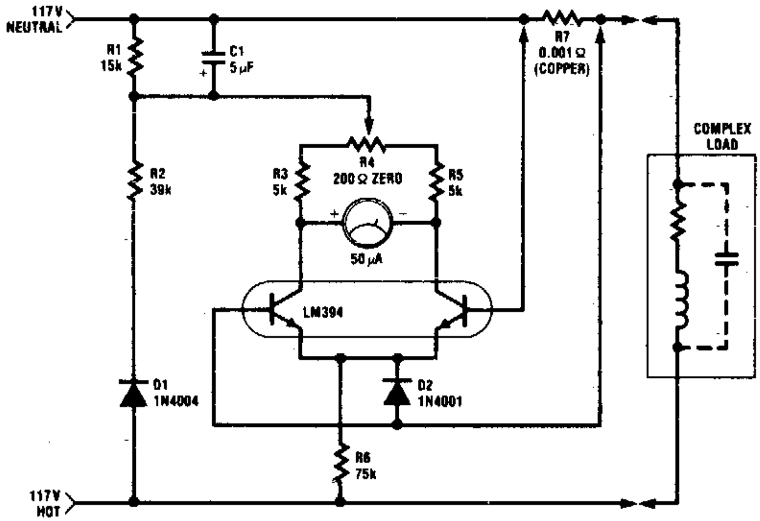
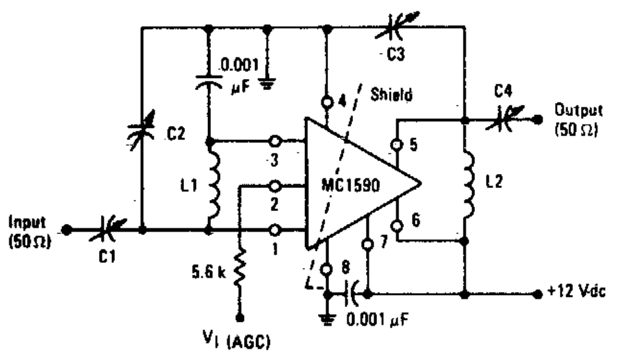


Fig. 63-4

Circuit Notes

The circuit is intended for 117 Vac ± 50 Vac operation, but can be easily modified for higher or lower voltages. It measures true (nonreactive) power being delivered to the load and requires no external power supply. Idling power drain is only 0.5 W. Load current sensing voltage is only 10 mV, keeping load voltage loss to 0.01%. Rejection of reactive load currents is better than 100:1 for linear loads. Nonlinearity is about 1% full scale when using a 50 µA meter movement.

60 MHz POWER GAIN TEST CIRCUIT



L1 = 7 Turns, #20 AWG Wire, 5/16" Dia., 5/8" Long
L2 = 6 Turns, #14 AWG Wire, 9/16" Dia., 3/4" Long

C1,C2,C3 = (1-30) pF
C4 = (1-10) pF

Fig. 63-5

64

Power Supplies (Fixed)

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SWITCHING REGULATOR OPERATING AT 200 kHz

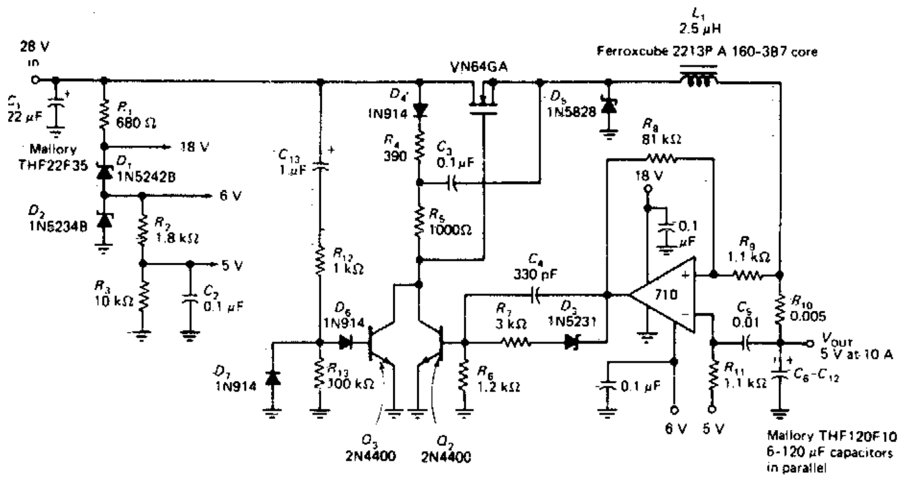


Fig. 64-1

Circuit Notes

This circuit provides a regulated dc with less than 100 mV of ripple for microprocessor applications. Necessary operating voltages are taken from the bleeder resistor network connected across the unregulated 28 V supply. The output of the LM710 comparator (actually an oscillator running at 200 kHz) is fed through a level-shifting circuit to the base of bipolar transistor Q2. This transistor is part of a bootstrap circuit necessary to turn the power MOSFET full on in totem-pole MOSFET arrays.

5 V, 0.5 A POWER SUPPLY

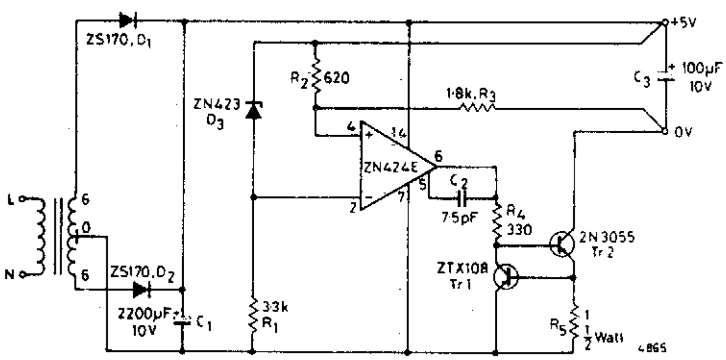
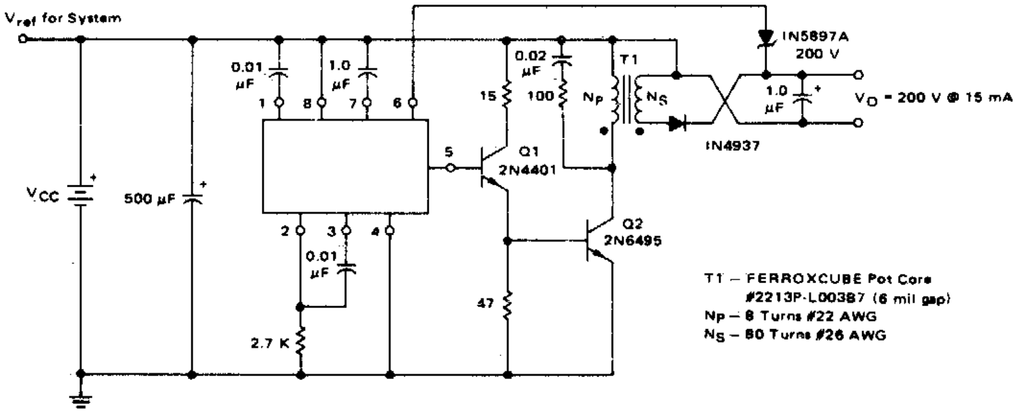


Fig. 64-2

Circuit Notes

The circuit is essentially a constant source modified by the feedback components R2 and R3 to give a constant voltage output. The output of the ZN424E need only be 2 volts above the negative rail, by placing the load in the collector of the output transistor Tr2. The current circuit is achieved by Tr1 and R5. This simple circuit has the following performance characteristics: Output noise and ripple (full load) = 1 mV rms. Load regulation (0 to 0.5 A) = 0.1%. Temperature coefficient = ± 100 ppm/°C. Current limit = 0.65 A.

3 W SWITCHING REGULATOR APPLICATION CIRCUIT



3-Watt Switching Regulator - converts 5 V to 200 V for gas discharge displays such as Burroughs Panaplex and Beckman.

Fig. 64-3

REGULATED SPLIT POWER SUPPLIES FROM A SINGLE SUPPLY

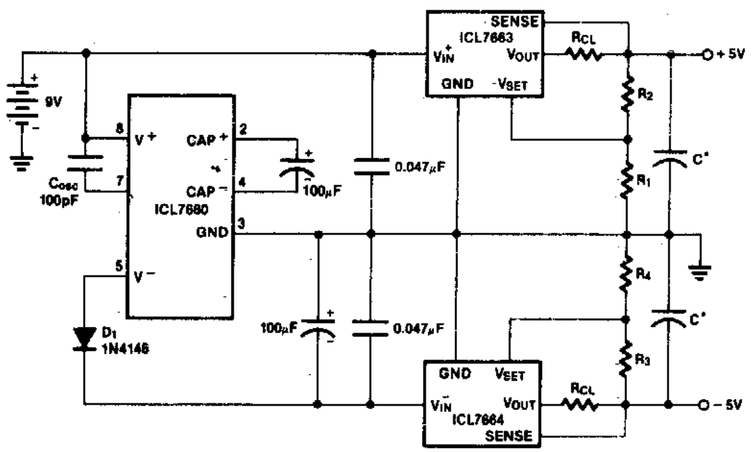


Fig. 64-4

*Values depend on load characteristics

Circuit Notes

The oscillation frequency of the ICL7660 is reduced by the external oscillator capacitor, so that it inverts the battery voltage more efficiently.

SWITCHING STEP-DOWN REGULATOR

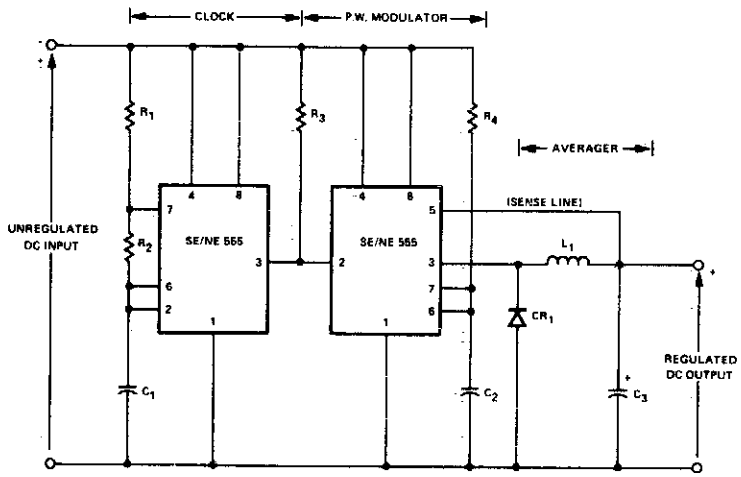


Fig. 64-5

SINGLE-ENDED REGULATOR

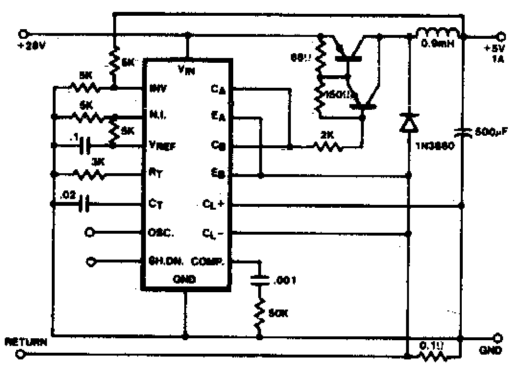


Fig. 64-6

Circuit Notes

In this conventional single-ended regulator circuit, the two outputs of the SG1524 are connected in parallel for effective 0-90% duty-cycle modulation. The use of an output inductor requires an RC phase compensation network for loop stability.

±50 V PUSH-PULL SWITCHED MODE CONVERTER

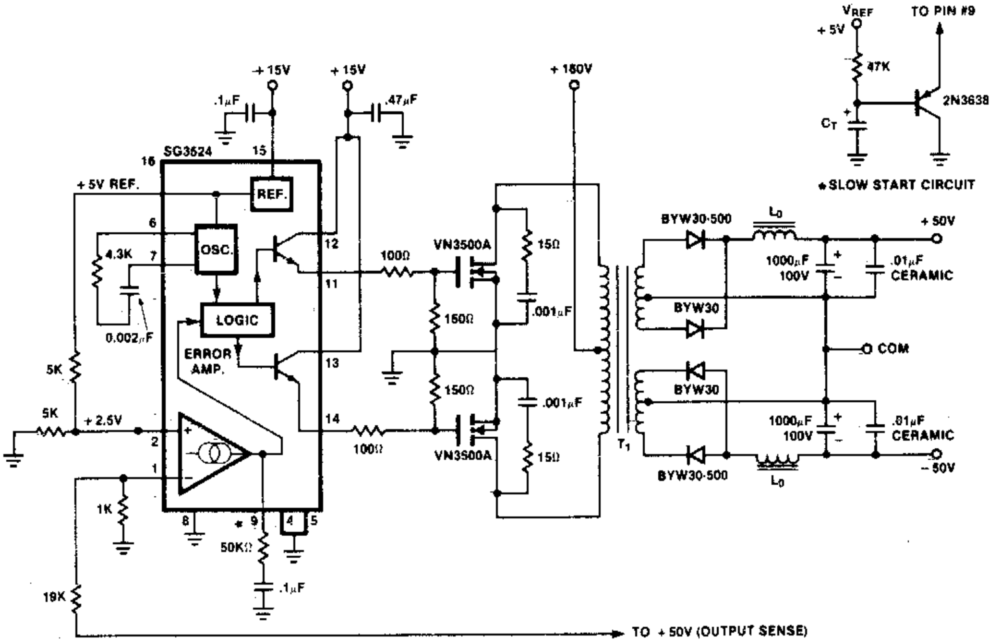


Fig. 64-7

5 V/0.5 A BUCK CONVERTER

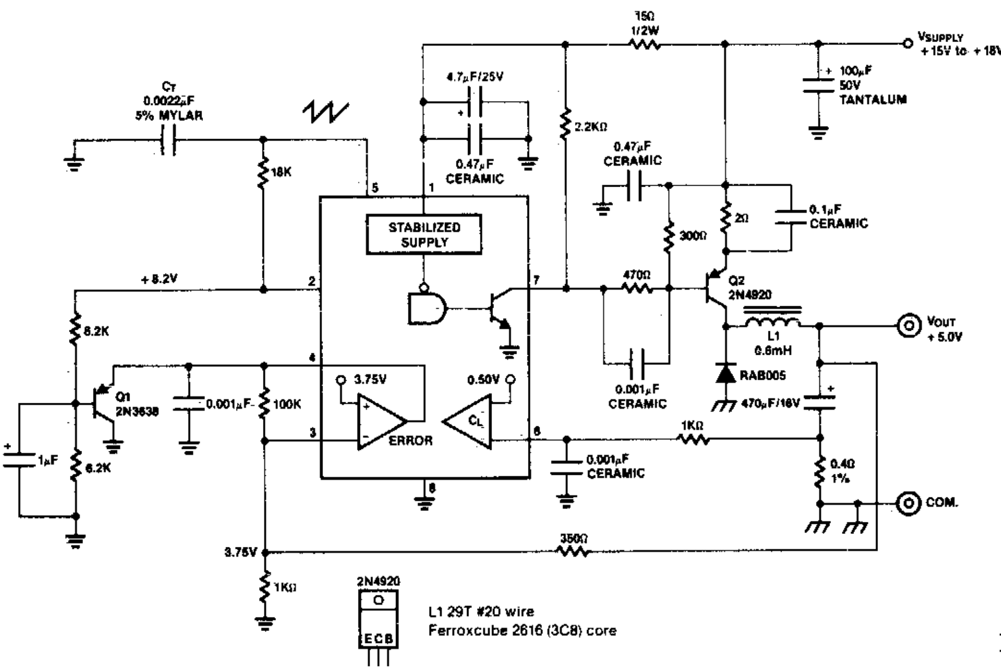


Fig. 64-8

±50 V FEED FORWARD SWITCH MODE CONVERTER

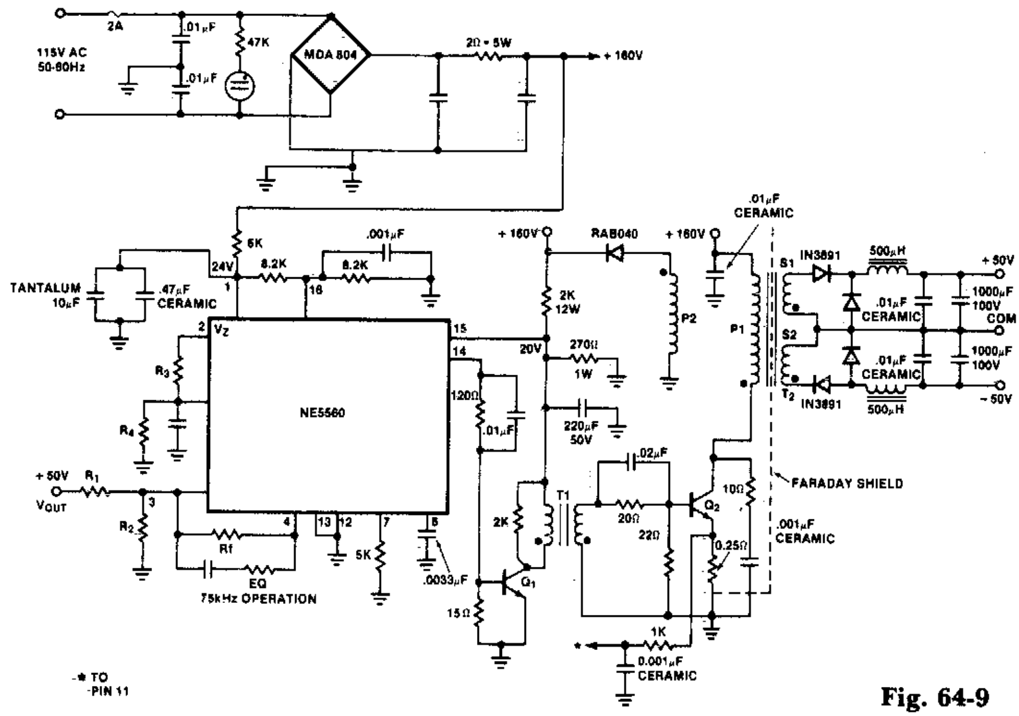


Fig. 64-9

TRAVELLER'S SHAVER ADAPTER

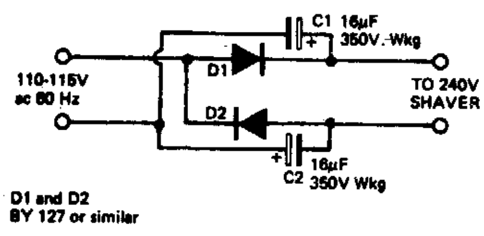


Fig. 64-10

Circuit Notes

Many countries have 115 volts mains supplies. This can be a problem if your electric shaver is designed for 220/240 volts only. This simple rectifier voltage doubler enables motor driven 240 volt shavers to be operated at full speed from a 115 volt supply. As the output voltage is dc, the circuit can only be used to drive small ac/dc motors. It cannot be used, for example, to operate vibrator-type shavers, or radios unless the latter are ac/dc operated.

100 Vrms VOLTAGE REGULATOR

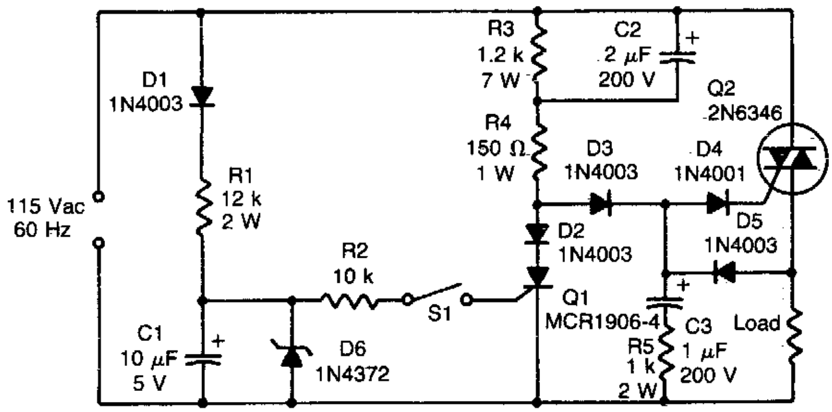


Fig. 64-11

TRANSISTOR INCREASES ZENER RATING

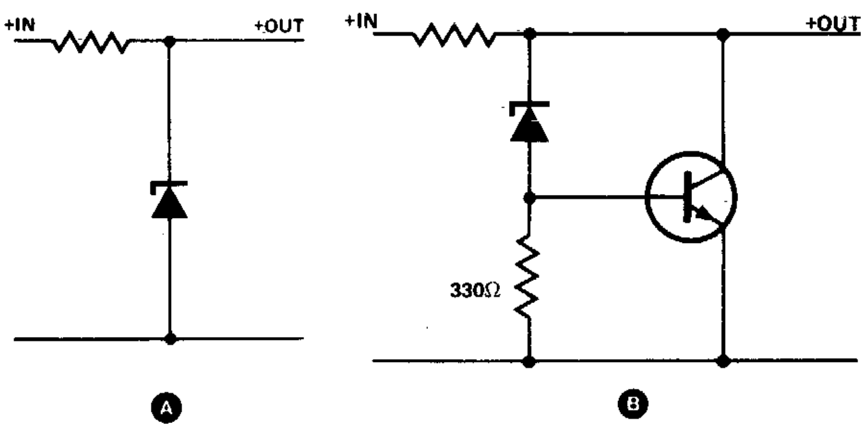
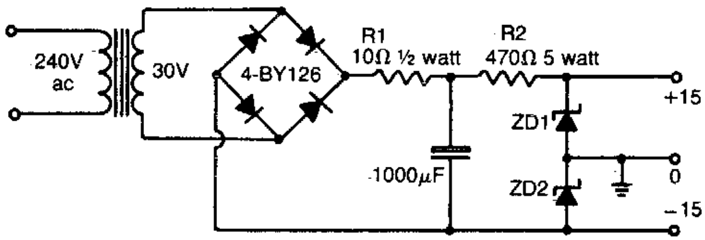


Fig. 64-12

Circuit Notes

The simple zener shunt in A may not handle sufficient current if the zener available is of low wattage. A power transistor will do most of the work for the zener as shown in B. Once the zener starts conducting, a bias voltage develops across the resistor (330 Ω to 1 K), turning on the transistor. The output voltage is 0.7 V greater than the zener voltage.

DUAL POLARITY POWER SUPPLY



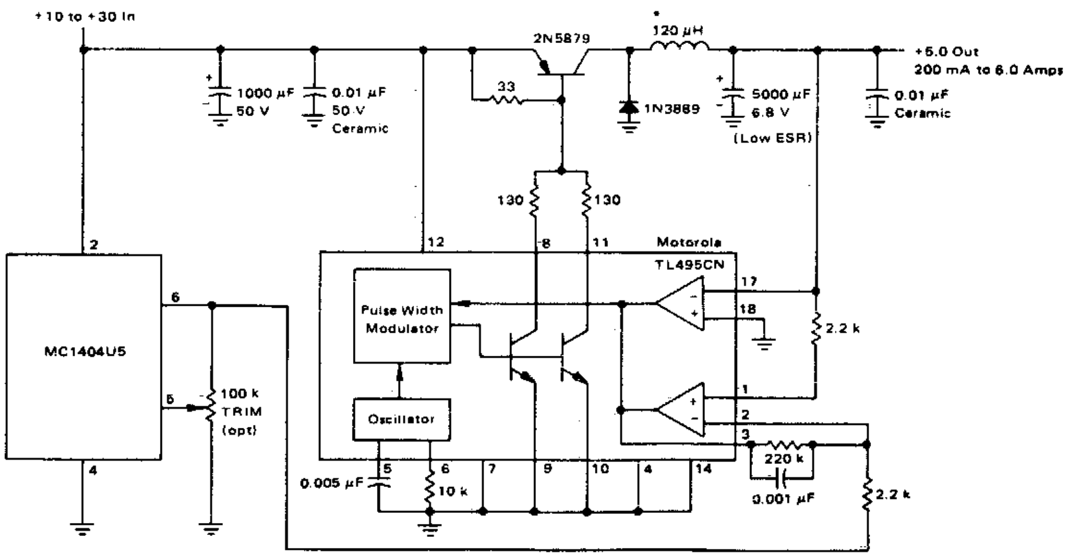
ZD 1 and ZD 2 = 15V ZENER DIODES—BZX70 C15

Fig. 64-13

Circuit Notes

This simple circuit gives a positive and negative supply from a single transformer winding and one full-wave bridge. Two zener diodes in series provide the voltage division and their centerpoint is grounded. (The filter capacitor must not be grounded via its case).

5.0 V/6.0 A 25 kHz SWITCHING REGULATOR WITH SEPARATE ULTRA-STABLE REFERENCE



*40 Turns #16 Wire, Arnold A-894075-2 Ferrite Core

Fig. 64-14

MOBILE VOLTAGE REGULATOR

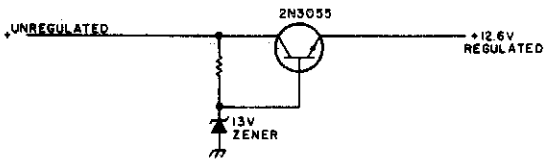
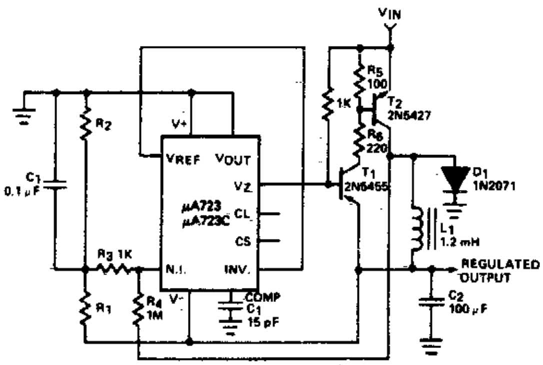


Fig. 64-15

Circuit Notes

This simple mobile voltage regulator circuit may save your two meter or CB transceiver if the voltage regulator fails. The 2N3055 should be heat sinked if current drawn by the rig is in excess of 2 A on transmit. This circuit will do little under normal operating conditions, but could save expensive equipment if the vehicle's electrical system loses regulation.

NEGATIVE SWITCHING REGULATOR

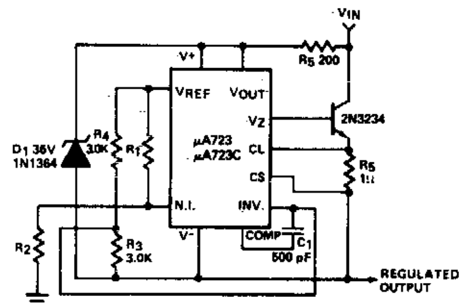


TYPICAL PERFORMANCE

Regulated Output Voltage	-15V
Line Regulation (ΔV_{IN} = 20V)	8 mV
Load Regulation (ΔI_L = 2A)	6 mV

Fig. 64-16

POSITIVE FLOATING REGULATOR

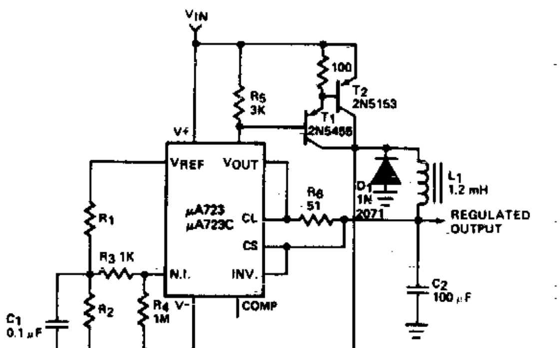


TYPICAL PERFORMANCE

Regulated Output Voltage	+50V
Line Regulation (ΔV_{IN} = 20V)	15 mV
Load Regulation (ΔI_L = 50 mA)	20 mV

Fig. 64-18

POSITIVE SWITCHING REGULATOR

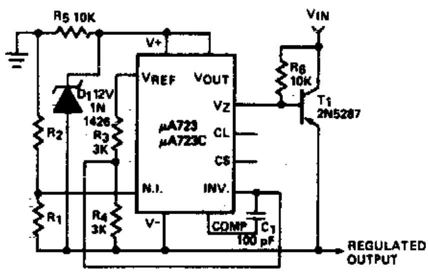


TYPICAL PERFORMANCE

Regulated Output Voltage	+5V
Line Regulation (ΔV_{IN} = 30V)	10 mV
Load Regulation (ΔI_L = 2A)	80 mV

Fig. 64-17

NEGATIVE FLOATING REGULATOR

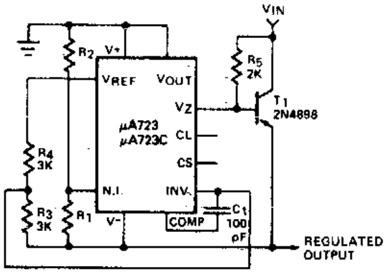


TYPICAL PERFORMANCE

Regulated Output Voltage	-100V
Line Regulation (ΔV_{IN} = 20V)	30 mV
Load Regulation (ΔI_L = 100 mA)	20 mV

Fig. 64-19

NEGATIVE VOLTAGE REGULATOR



TYPICAL PERFORMANCE

Regulated Ouput Voltage	-15V
Line Regulation (ΔV_{IN} = 3V)	1 mV
Load Regulation (ΔI_L = 100 mA)	2 mV

Fig. 64-20

HIGH STABILITY 10 V REGULATOR

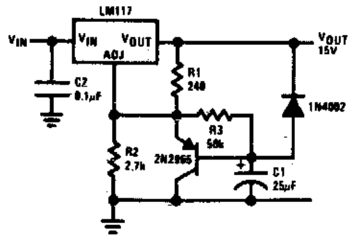


Fig. 64-23

–15 V NEGATIVE REGULATOR

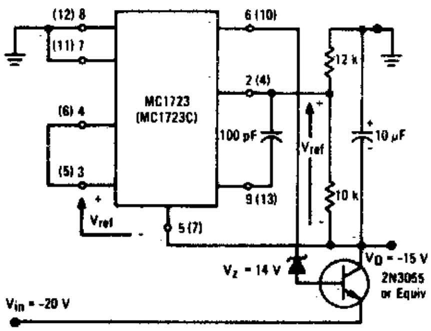


Fig . 64-21

5 V/1 A SWITCHING REGULATOR

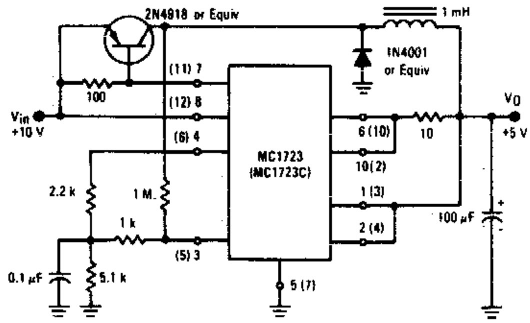


Fig. 64-24

SLOW TURN-ON 15 V REGULATOR

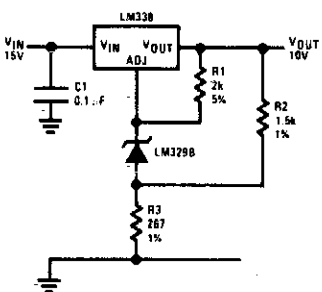


Fig. 64-22

15 V/1 A REGULATOR
WITH REMOTE SENSE

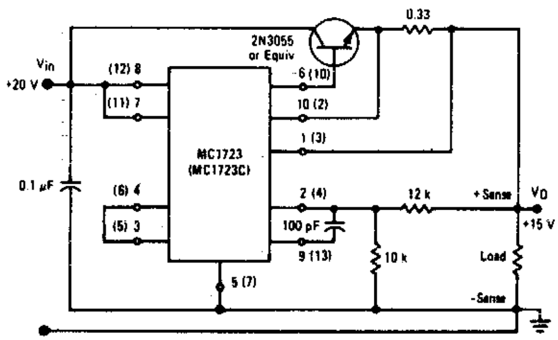


Fig. 64-25

LOW RIPPLE POWER SUPPLY

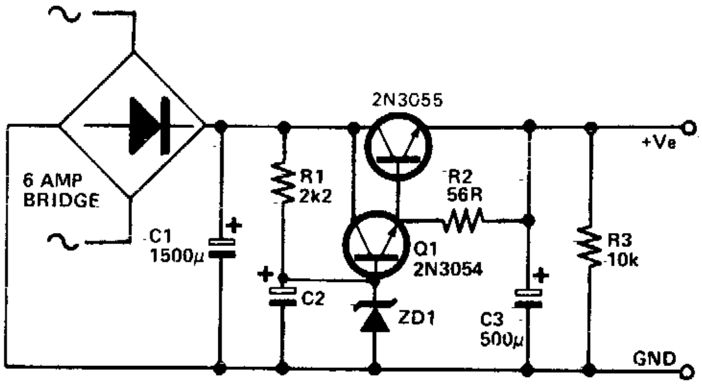


Fig. 64-26

Circuit Notes

This circuit may be used where a high current is required with a low ripple voltage (such as in a high powered class AB amplifier when high quality reproduction is necessary). Q1, Q2, and R2 may be regarded as a power darlington transistor. ZD1 and R1 provide a reference voltage at the base of Q1. ZD1 should be chosen thus: ZD1 = $V_{out} - 1.2$. C2 can be chosen for the degree of smoothness as its value is effectively multiplied by the combined gains of Q1/Q2, if 100 μF is chosen for C2, assuming minimum hfe for Q1 and Q2, C = 100 × 15(Q1) × 25(Q2) = 37,000 μF.

5.0 V/10 A REGULATOR

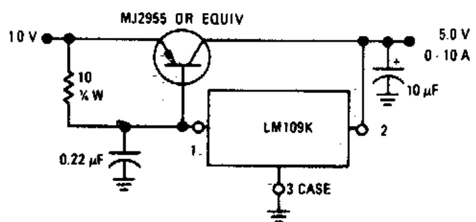


Fig. 64-27

5.0 V/3.0 A REGULATOR

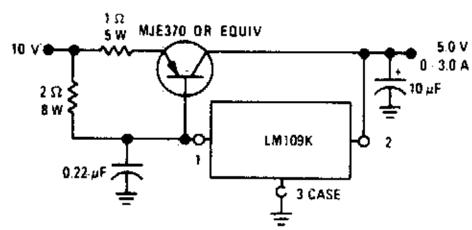
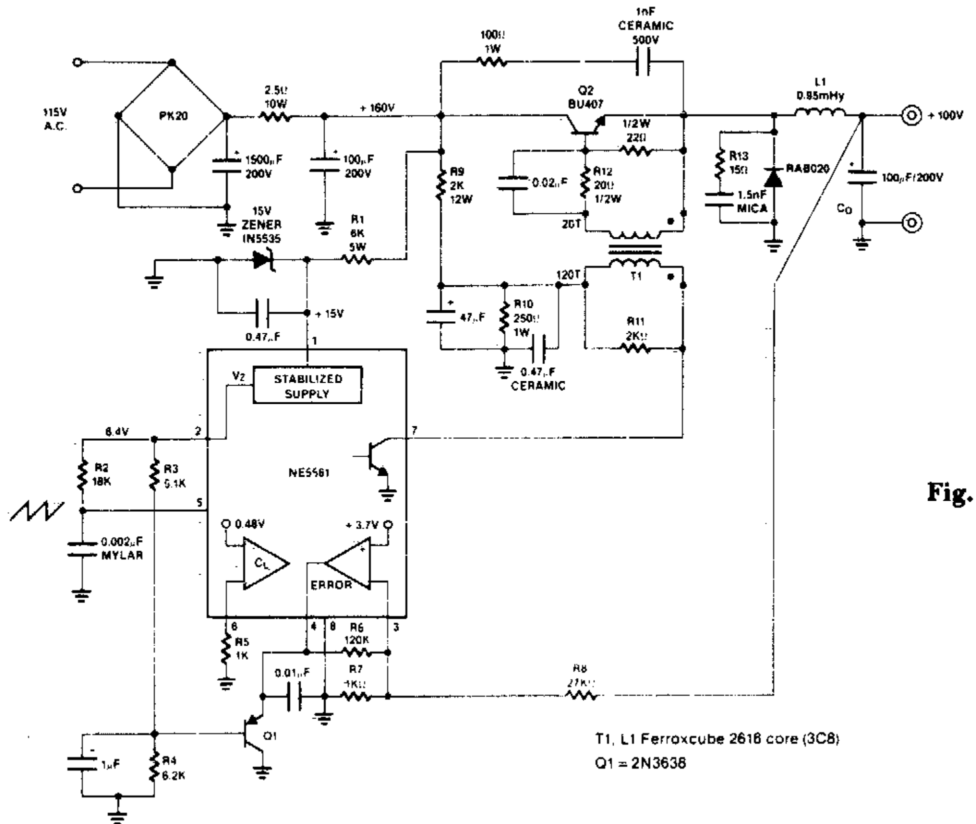


Fig. 64-28

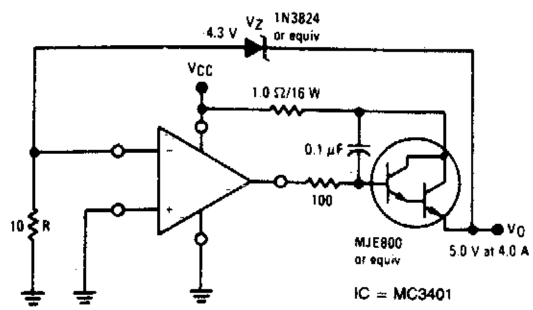
100 V/0.25 A SWITCH MODE CONVERTER



T1, L1 Ferroxcube 2616 core (3C8)
Q1 = 2N3638

Fig. 64-29

VOLTAGE REGULATOR

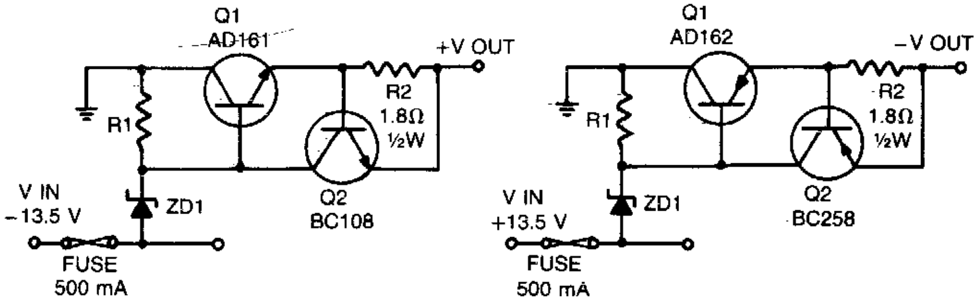


IC = MC3401

$V_O = V_Z + 0.6$ Vdc
NOTE 1: R is used to bias the zener.
NOTE 2: If the Zener TC is positive, and equal in magnitude to the negative TC of the input to the operational amplifier (≈ 2.0 mV/°C), the output is zero-TC. A 7.0-Volt Zener will give approximately zero-TC.

Fig. 64-30

LOW VOLTAGE REGULATORS WITH SHORT CIRCUIT PROTECTION



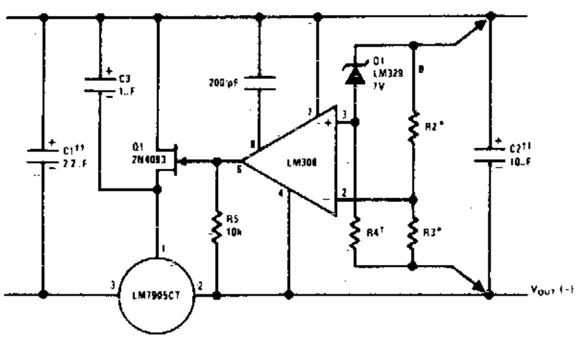
VOLTAGE	ZD1 400mW	R1
6V	6V2	680Ω
7.5V	7V5	390Ω
9V	9V1	220Ω

Fig. 64-31

Circuit Notes

These short-circuit protected regulators give 6, 7.5, and 9 V from an automobile battery supply of 13.5 V nominal; however, they will function just as well if connected to a smoothed dc output from a transformer/rectifier circuit. Two types are shown for both positive and negative ground systems. The power transistors can be mounted on the heatsink without a mica insulating spacer thus allowing for greater cooling efficiency. Both circuits are protected against overload or short-circuits. The current cannot exceed 330 mA. Under normal operating conditions the voltage across R2 does not rise above the 500 mV necessary to turn Q2 on and the circuit behaves as if there was only Q1 present. If excessive current is drawn, Q2 turns on and cuts off Q1, protecting the regulating transistor. The table gives the values of R1 for different zener voltages.

HIGH STABILITY 1 A REGULATOR



Load and line regulation < 0.01% temperature stability ≤ 0.2%

† Determines Zener current

†† Solid tantalum

* Select resistors to set output voltage. 2 ppm/°C tracking suggested

Fig. 64-32

100 V/0.25 A SWITCH MODE CONVERTER

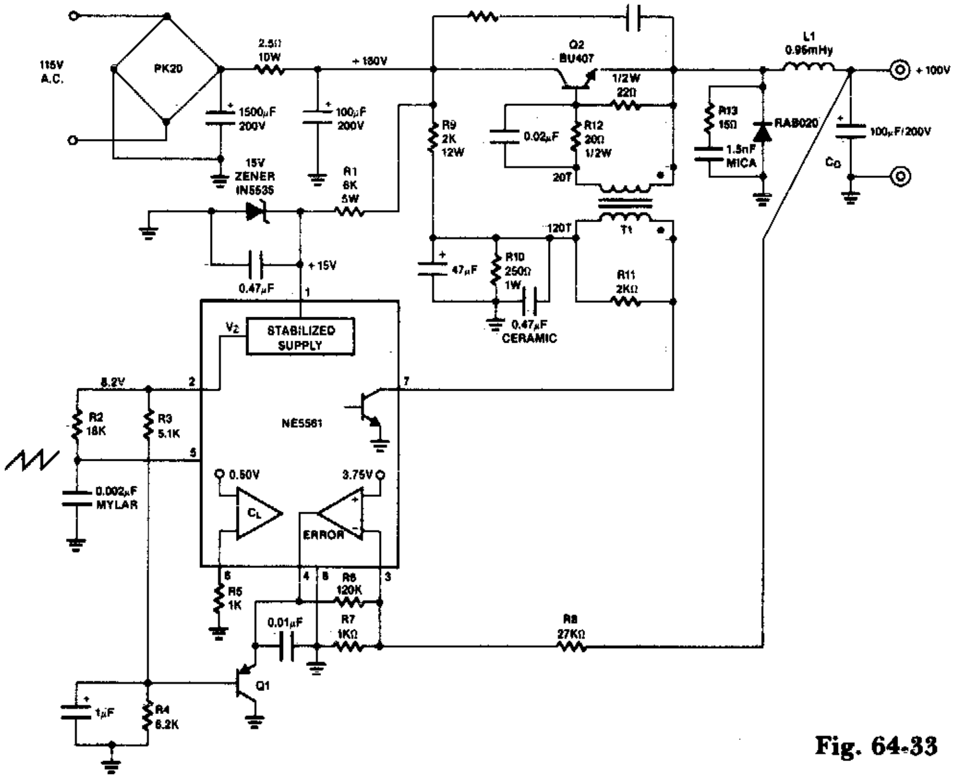


Fig. 64-33

65

Power Supplies (Variable)

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DUAL OUTPUT BENCH POWER SUPPLY

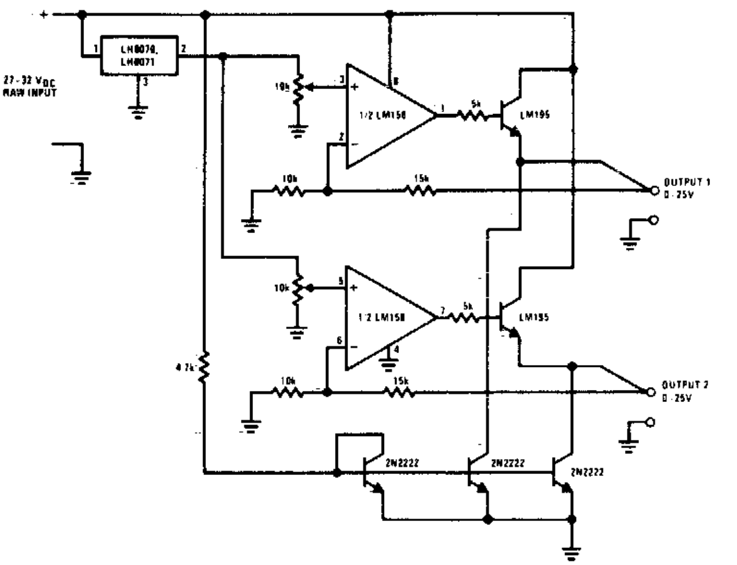
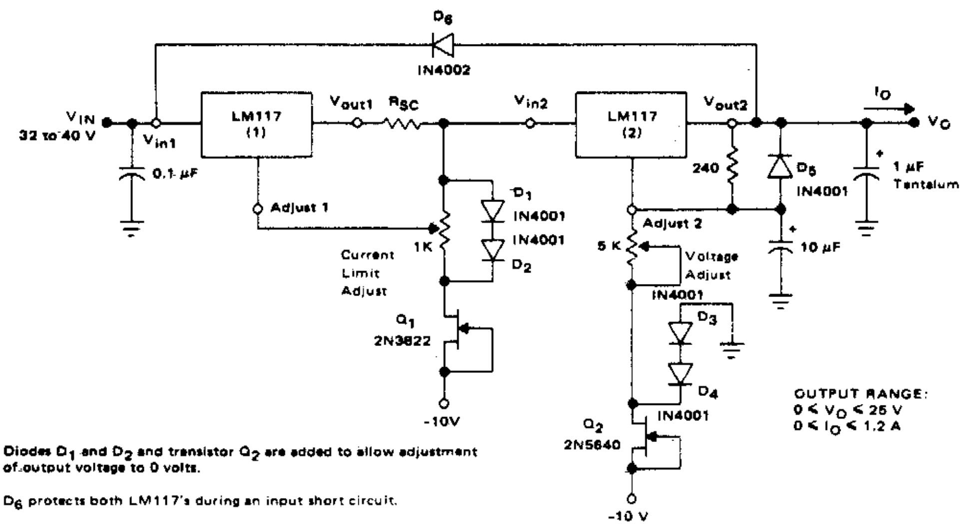


Fig. 65-1

POWER SUPPLY WITH ADJUSTABLE CURRENT LIMIT AND OUTPUT VOLTAGE

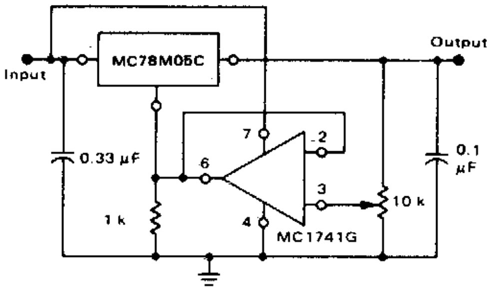


Diodes D_1 and D_2 and transistor Q_2 are added to allow adjustment of output voltage to 0 volts.

D_6 protects both LM117's during an input short circuit.

Fig. 65-2

ADJUSTABLE OUTPUT REGULATOR



V_O, 7.0 V to 20 V
$V_{IN} - V_O \geq 2.0$ V

Fig. 65-3

Circuit Notes

The addition of an operational amplifier allows adjustment to higher or intermediate values while retaining regulation characteristics. The minimum voltage obtainable with this arrangement is 2.0 volts greater than the regulator voltage.

RF PROBE FOR VTVM

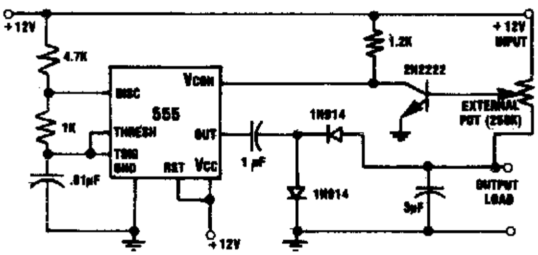


Fig. 65-4

Circuit Notes

This circuit combines a 555 timer with a 2N2222 transistor and an external potentiometer. The pot adjusts the output voltage to the desired value. To regulate the output voltage, the 2N2222 varies the control voltage of the 555 IC, increasing or decreasing the pulse repetition rate. A 1.2 K resistor is used as a collector load. The transistor base is driven from the external pot. If the output voltage becomes less negative, the control voltage moves closer to ground, causing the repetition rate of the 555 to increase, which, in turn, causes the 3 μF capacitor to charge more frequently. Output voltage for the circuit is 0 to 10 V, adjusted by the external pot. Output regulation is less than five percent for 0 to 10 mA and less than .05 percent for 0 to 0.2 mA.

REGULATED VOLTAGE DIVIDER

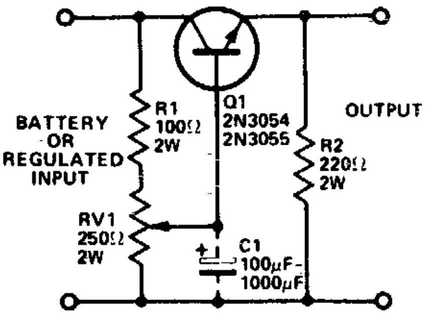


Fig. 65-5

Circuit Notes

ICs requiring 3.6 or 6 volts can be run from a battery or fixed regulated supply of a higher voltage by using the circuit shown. The transistor should be mounted on a heatsink as considerable power will be dissipated by its collector. Additional filtering can be obtained by fitting a capacitor (C1) as shown. The capacitance is effectively multiplied by the gain of the transistor. A ripple of 200 mV (peak to peak) at the input can be reduced to 2 mV in this fashion. Maximum output current depends on the supply rating and transistor type (with heatsink) used.

VARIABLE ZENER DIODE

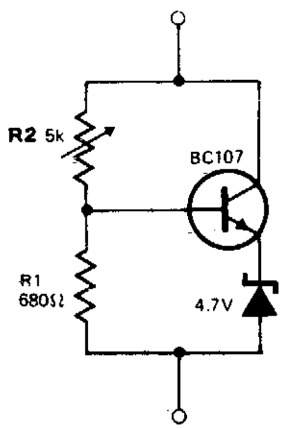
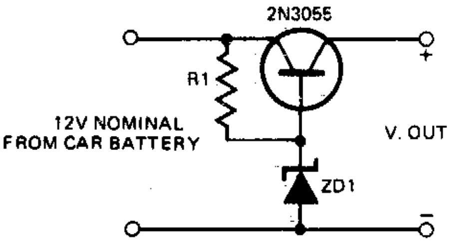


Fig. 65-6

Circuit Notes

The circuit behaves like a zener diode over a large range of voltages. The current passing through the voltage divider R1-R2 is substantially larger than the transistor base current and is in the region of 8 mA. The stabilizing voltage is adjustable over the range 5-45 V by changing the value of R2. The total current drawn by the circuit is variable over the range 15 mA to 50 mA. This value is determined by the maximum dissipation of the zener diode. In the case of a 250 mW device, this is of the order of 50 mA.

12 V TO 9, 7.5 or 6 V CONVERTER



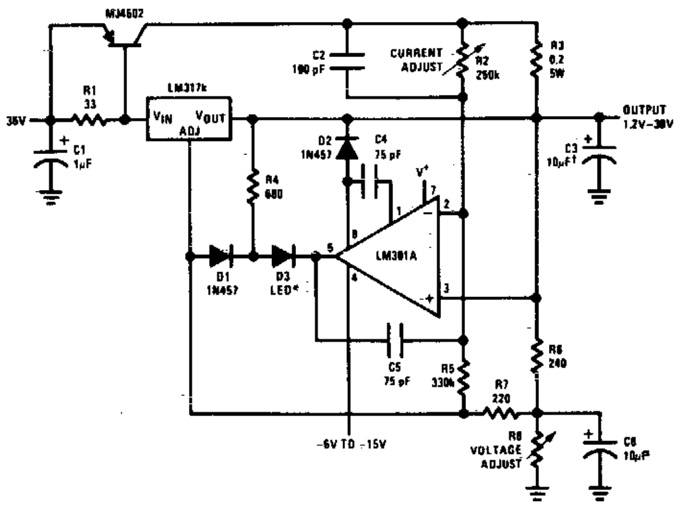
OUTPUT VOLTAGE	9	7.5	6
R1 (½ WATT)	180Ω	270Ω	330Ω
ZENER DIODE (250 mW)	10V	8V1	6V6

Circuit Notes

This circuit enables transistorized items such as radio, cassettes, and other electrical devices to be operated from a car's electrical supply. The table gives values for resistors and specified diode types for different voltage. Should more than one voltage be required a switching arrangement could be incorporated. For high currents, the transistor should be mounted on a heatsink.

Fig. 65-7

5 A CONSTANT VOLTAGE/CONSTANT CURRENT REGULATOR



†Solid tantalum

*Lights in constant current mode

Fig. 65-8

POWER PACK FOR BATTERY-POWERED CALCULATORS, RADIOS, OR CASSETTE PLAYERS

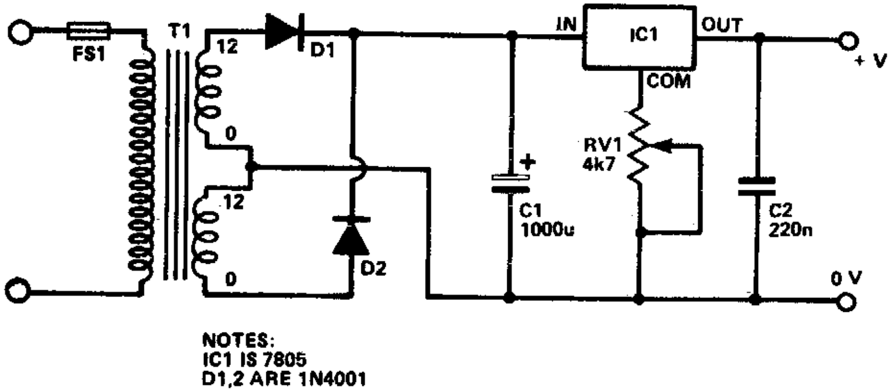


Fig. 65-9

Circuit Notes

This circuit gives a regulated output of between 5 V and 15 Vdc, adjusted and set by a preset resistor. Current output up to about 350 mA. An integrated circuit regulates the output voltage and although this IC (the 7805) is normally used in a fixed-voltage (5 Vdc) supply it is for a variable output voltage.

PRECISION HIGH VOLTAGE REGULATOR

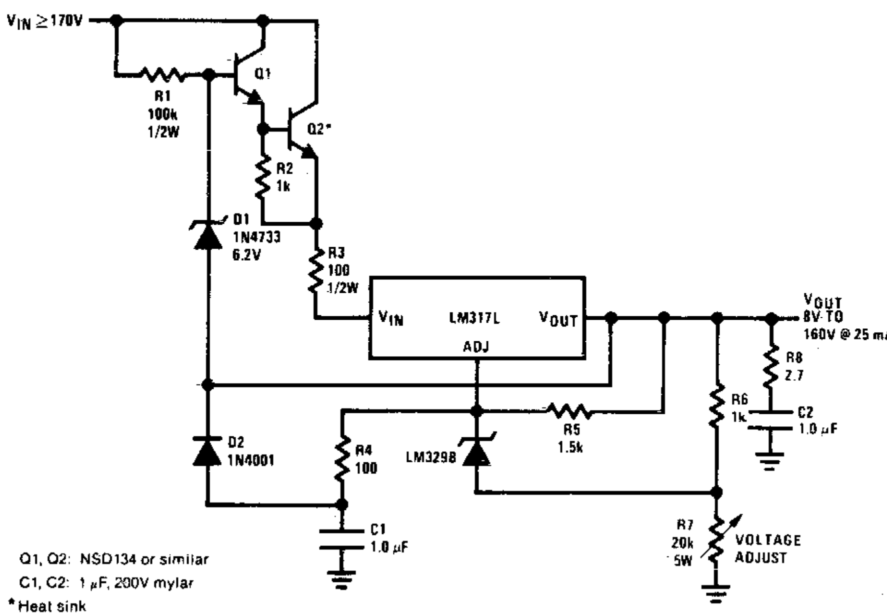
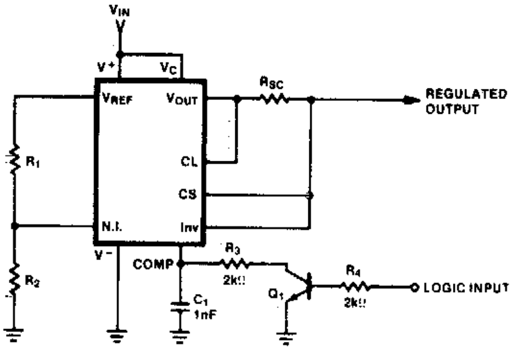


Fig. 65-10

REMOTE SHUTDOWN REGULATOR WITH CURRENT LIMITING (V_{out} = 2 TO 7 V)



IC = μA723

$$V_{OUT} = \left[V_{REF} \times \frac{R_2}{R_1 + R_2}\right]$$

Fig. 65-11

0 TO 30 V REGULATOR

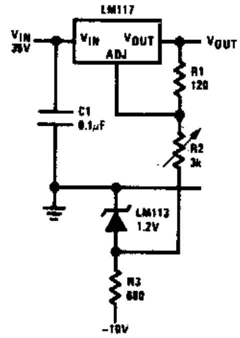
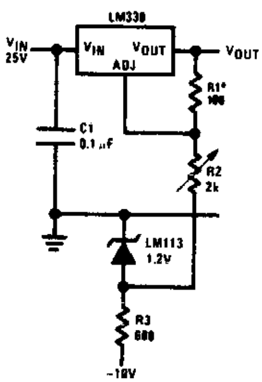


Fig. 65-13

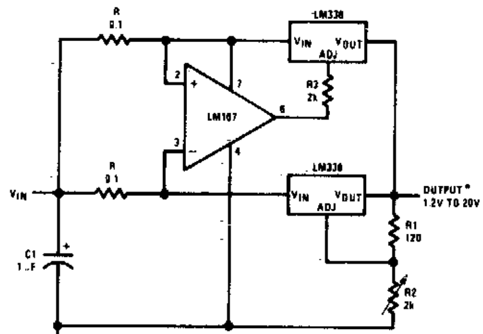
0 TO 22 V REGULATOR



*R1=240Ω, R2 = 5k for LM138 and LM238

Fig. 65-12

10 A REGULATOR



*Minimum load—100 mA

Fig. 65-14

ADJUSTABLE REGULATOR 0-10 V AT 3 A

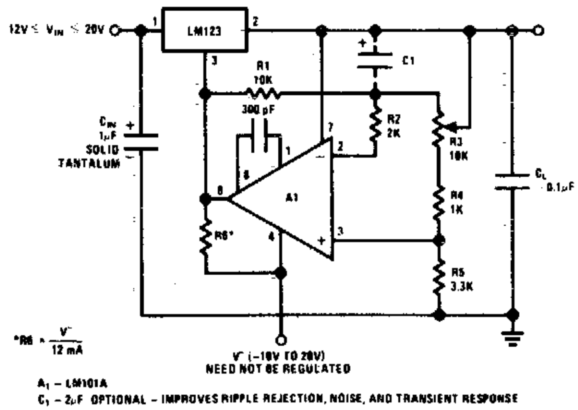
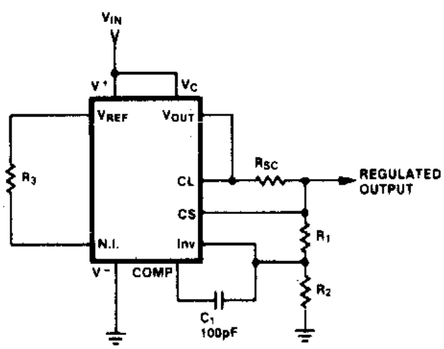


Fig. 65-15

HIGH VOLTAGE REGULATOR (V_{out} = + 7 V TO 37 V)



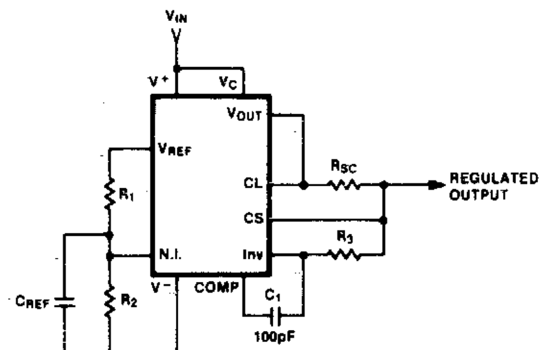
$$V_{OUT} = \left[V_{REF} \times \frac{R_1 + R_2}{R_2}\right]$$

$R_3 = \frac{R_1 R_2}{R_1 + R_2}$ for minimum temperature drift

R3 may be eliminated for minimum component count

Fig. 65-16

LOW VOLTAGE REGULATOR (V_{out} = 2 TO 7 V)



$$V_{OUT} = \left[V_{REF} \times \frac{R_2}{R_1 + R_2}\right]$$

$R_3 = \frac{R_1 R_2}{R_1 + R_2}$ for minimum temperature drift

Fig. 65-17

SIMPLE SPLIT POWER SUPPLY

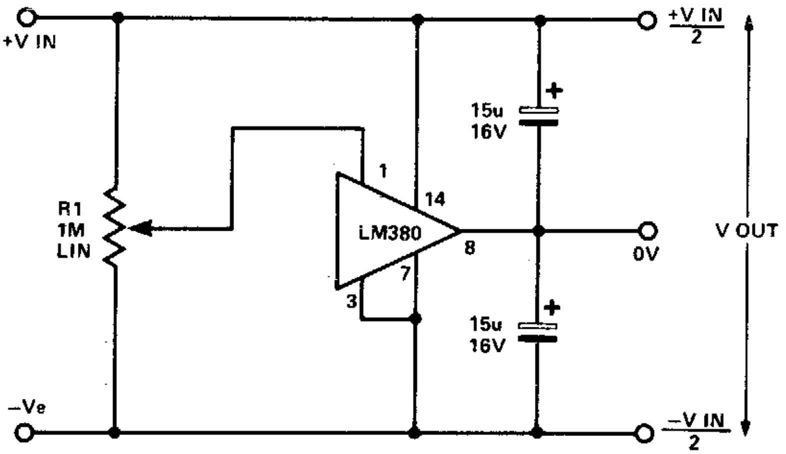


Fig. 65-18

Circuit Notes

This circuit utilizes the quasi-complementary output stage of the popular LM380 audio power IC. The device is internally biased so that with no input the output is held midway between the supply rails. R1, which should be initially set to mid-travel, is used to nullify any inbalance in the output. Regulation of V_{out} depends upon the circuit feeding the LM380, but positive and negative outputs will track accurately irrespective of input regulation and unbalanced loads. The free-air dissipation is a little over 1 watt, and so extra cooling may be required. The device is fully protected and will go into thermal shutdown if its rated dissipation is exceeded. Current limiting occurs if the output current exceeds 1.3 A. The input voltage should not exceed 20 V.

ADJUSTABLE OUTPUT REGULATOR

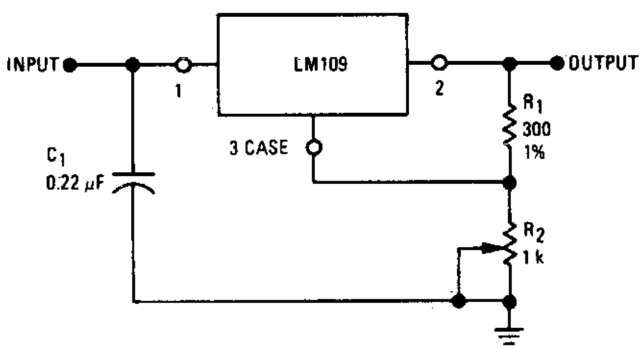
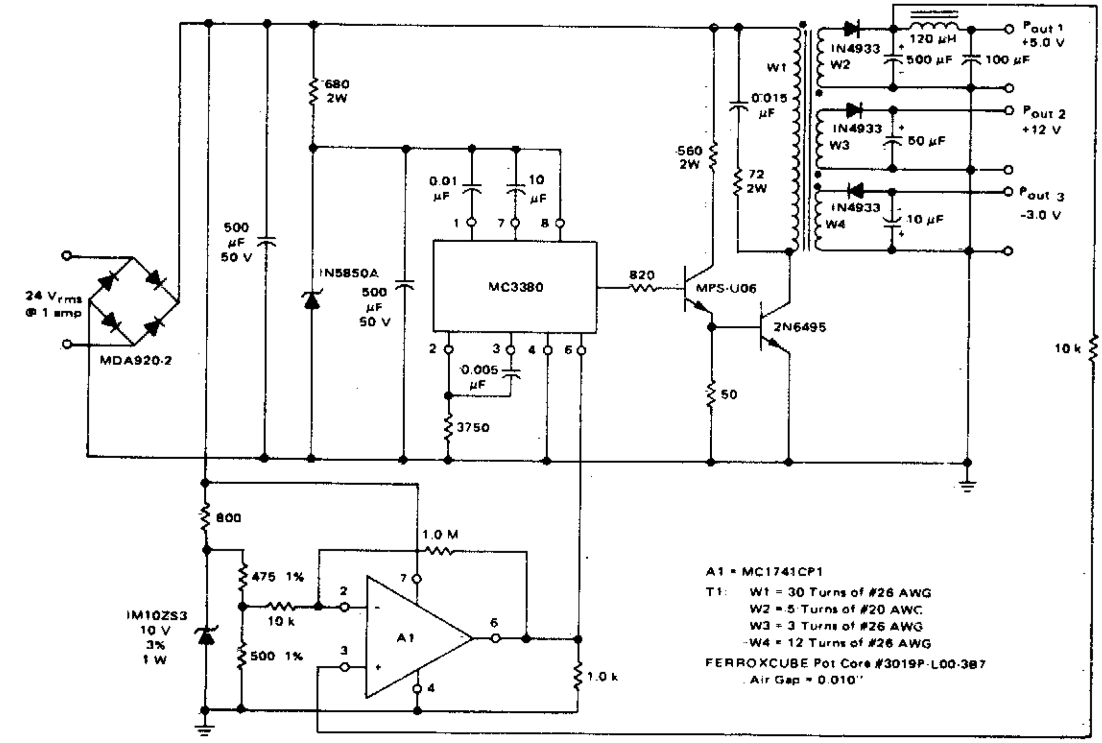


Fig. 65-19

MULTIPLE OUTPUT SWITCHING REGULATOR FOR USE WITH MPUs

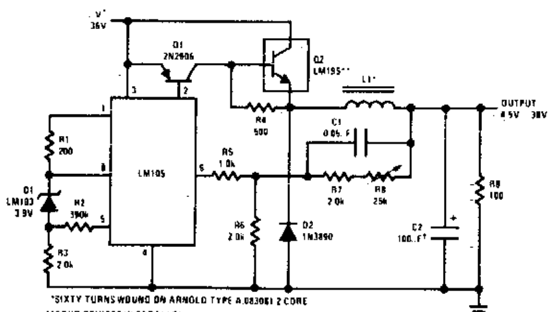


A1 = MC3380P

TYPICAL PERFORMANCE
P_{out1} = 4 Watts
(V_O = 5 V ± 5%)
5 V Ripple Component = 50 mV
(120 Hz + 20 kHz)
P_{out2} = 600 mW
(V_O = 12 V ± 10%)
P_{out3} = 3 mW
(V_O = -3 V ± 10%)

Fig. 65-20

6.0 A VARIABLE OUTPUT SWITCHING REGULATOR



*SIXTY TURNS WOUND ON ARNOLD TYPE A-083081-2 CORE
**FOUR DEVICES IN PARALLEL
†SOLID TANTALUM

Fig. 65-21

66

Power Supply Protection Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ELECTRONIC CROWBAR FOR AC OR DC LINES

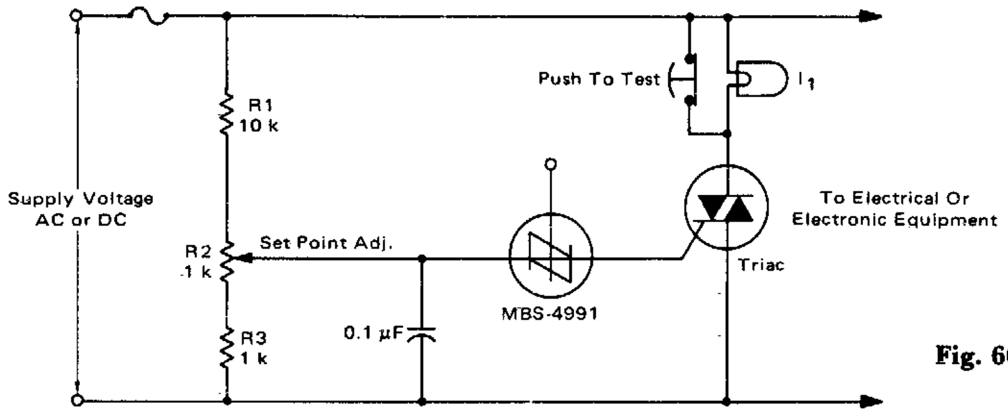


Fig. 66-1

Circuit Notes

For positive protection of electrical or electronic equipment, use this against excessive supply voltage. Due to improper switching, wiring, short circuits, or failure of regulators, an electronic crowbar circuit can quickly place a short circuit across the power lines, thereby dropping the voltage across the protected device to near zero and blowing a fuse. The triac and SBS are both bilateral devices, the circuit is equally useful on ac or dc supply lines. With the values shown for R1, R2, and R3, the crowbar operating point can be adjusted over the range of 60 to 120 volts dc or 42 to 84 volts ac. The resistor values can be changed to cover a different range of supply voltages. The voltage rating of the triac must be greater than the highest operating point as set by R2. I1 is a low power incandescent lamp with a voltage rating equal to the supply voltage. It may be used to check the set point and operation of the unit by opening the test switch and adjusting the input or set point to fire the SBS. An alarm unit such as the Mallory Sonalert may be connected across the fuse to provide an audible indication of crowbar action. (This circuit may not act on short, infrequent power line transients).

POWER PROTECTION CIRCUIT

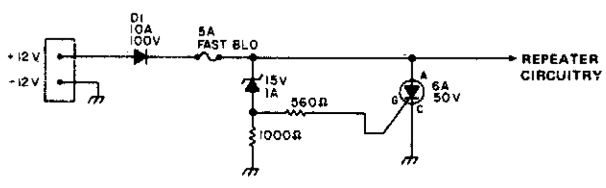


Fig. 66-2

Circuit Notes

To safeguard portable, emergency power repeaters from reverse or excessive voltage, D1 prevents incorrect polarity damage, and zener voltage determines the maximum voltage that will reach the rest of the circuitry. Use fast blowing fuse rated greater than the SCR current rating.

SIMPLE CROWBAR

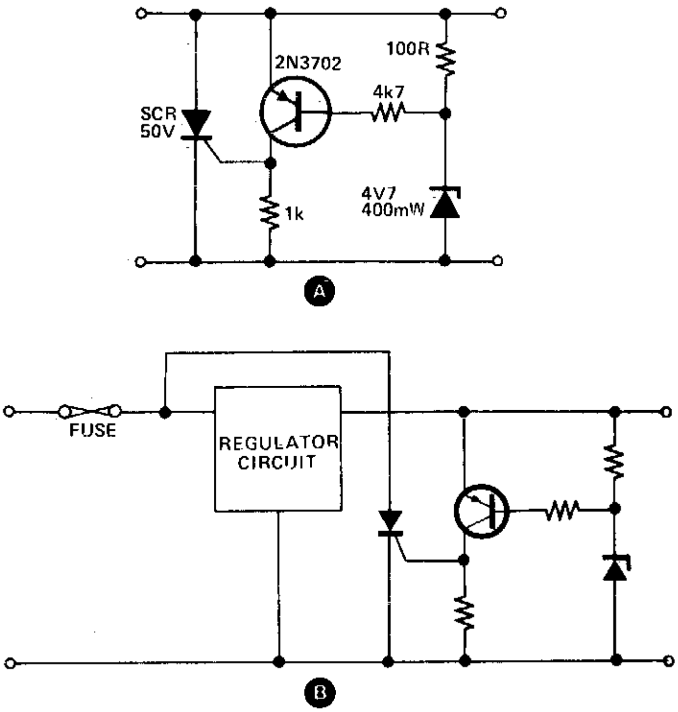


Fig. 66-3

Circuit Notes

These circuits provide overvoltage protection in case of voltage regulator failure or application of an external voltage. Intended to be used with a supply offering some form of short circuit protection, either foldback, current limiting, or a simple fuse. The most likely application is a 5 V logic supply, since TTL is easily damaged by excess voltage. The values chosen in A are for a 5 V supply, although any supply up to about 25 V can be protected by simply choosing the appropriate zener diode. When the supply voltage exceeds the zener voltage +0.7 V, the transistor turns on and fires the thyristor. This shorts out the supply, and prevents the voltage rising any further. In the case of a supply with only fuse protection, it is better to connect the thyristor the regulator circuit when the crowbar operates. The thyristor should have a current rating about twice the expected short circuit current and a maximum voltage greater than the supply voltage. The circuit can be reset by either switching off the supply, or by breaking the thyristor circuit with a switch.

OVERVOLTAGE PROTECTION WITH AUTOMATIC RESET

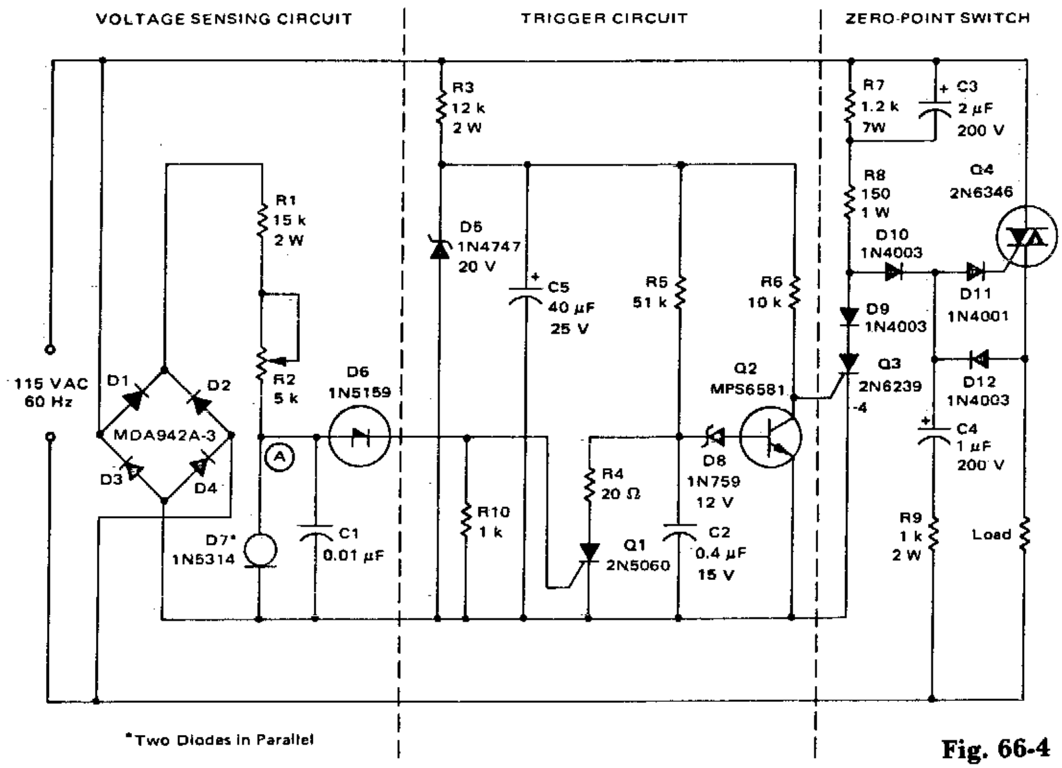


Fig. 66-4

OVERVOLTAGE PROTECTION FOR LOGIC

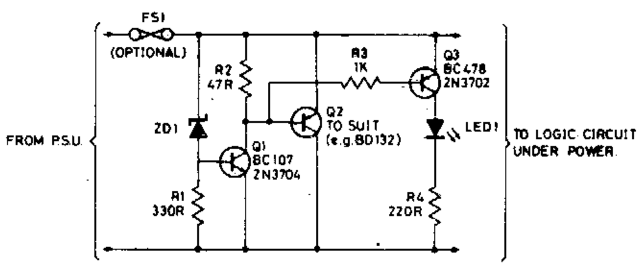


Fig. 66-5

Circuit Notes

Zener diode ZD1 senses the supply, and should the supply rise above 6 V, Q1 will turn on. In turn, Q2 conducts clamping the rail. Subsequent events depend on the source supply. It will either shut down, go into current limit or blow its supply fuse. None of these will damage the TTL chips. The rating of Q2 depends on the source supply, and whether it will be required to operate continuously in the event of failure. Its current rating has to be in excess of the source supply.

FAST ACTING POWER SUPPLY PROTECTION

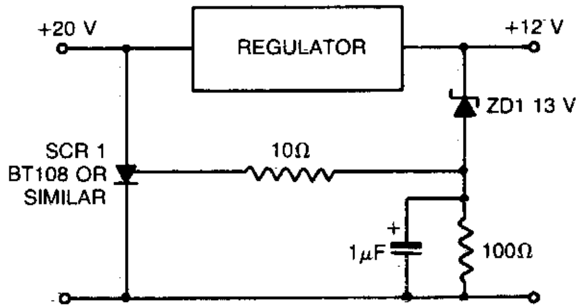


Fig. 66-6

Circuit Notes

When using a regulated power supply to reduce a supply voltage, there is always the danger that component failure in the power supply might lead to a severe overvoltage condition across the load. To cope with overvoltage situations, the circuit is designed to protect the load under overvoltage conditions. Component values given are for a 20 V supply with regulated output at 12 V. The zener diode can be changed according to whatever voltage is to be the maximum. If the voltage at the regulator output rises to 13 V or above, the zener diode breaks down and triggers the thyristor which shorts out the supply line and blows the main fuse.

5 V CROWBAR

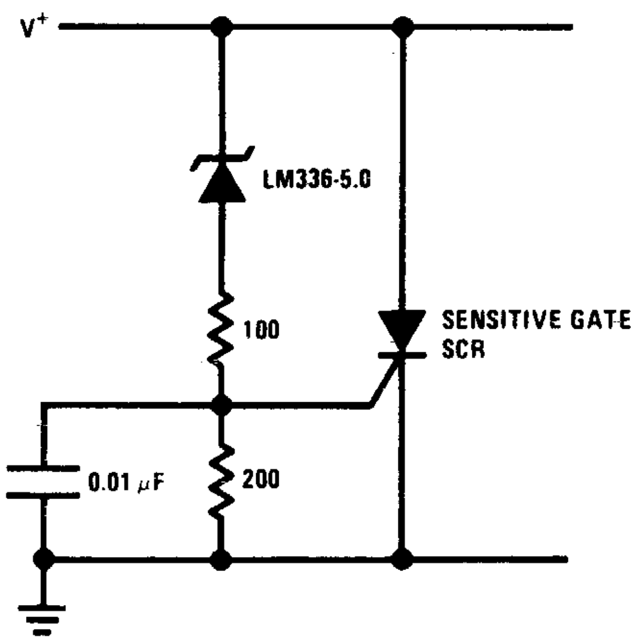


Fig. 66-7

67

Probes

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Logic Probe Yields Three Discrete States
Signal Injector/Tracer
Injector/Tracer
CMOS Logic Probe
RF Probe for VOM
100 K Megohm DC Probe
Audible TTL Probe
Logic Probe
Logic Test Probe with Memory
Logic Probe
Simple Logic Probe
Audio-RF Signal Tracer Probe
TTL Logic Tester

LOGIC PROBE YIELDS THREE DISCRETE STATES

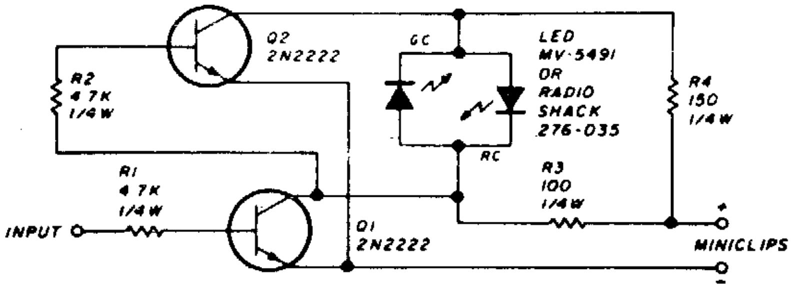
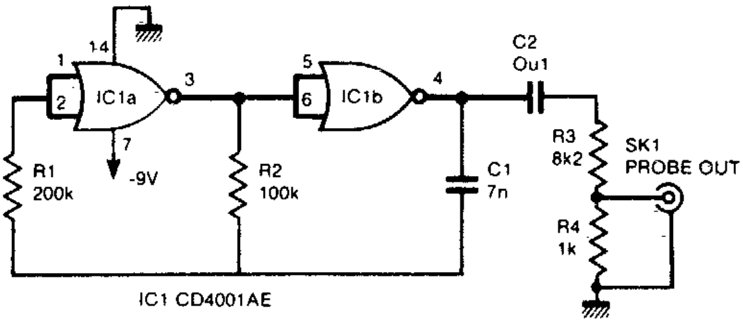


Fig. 67-1

Circuit Notes

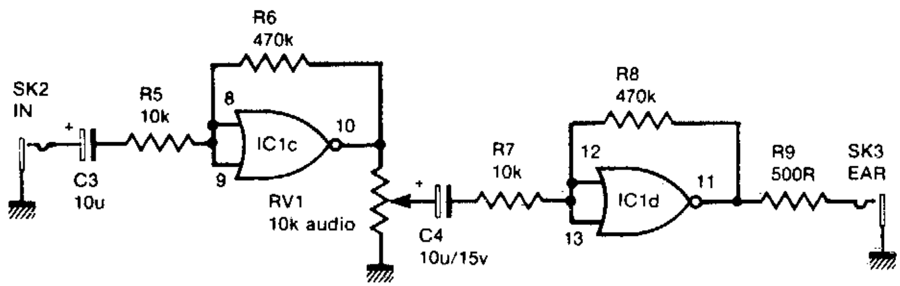
The circuit uses a dual LED. When power is applied to the probe through the power leads, and the input is touched to a low level or ground, Q1 is cut off. This will cause Q2 to conduct since the base is positive with respect to the emitter. With Q1 cut off and Q2 conducting, the green diode of the dual LED will be forward biased, yielding a green output. Touching the probe tip to a high level will cause Q1 and Q2 to complement, and the red diode will be forward biased, yielding a red output from the LED. An alternating signal will cause alternating conduction of the red and green diodes and will yield an indication approximately amber. In this manner, both static and dynamic signals can be traced with the logic probe.

SIGNAL INJECTOR/TRACER



Injector circuit diagram.

Fig. 67-2



Tracer circuit diagram.

Circuit Notes

The injector is a CMOS oscillator with period approximately equal to 1.4 × C1 × R2 seconds. The values are given for 1 kHz operation. Resistors R3 and R4 divide the output to 1 V. Whereas the oscillator employs the gates in their digital mode, the tracer used them in a linear fashion by applying negative feedback from output to input. They are used in much the same way as op amps. The circuit uses positive ground. It offers an advantage at the earphone output because one side of the earphone must be connected to ground via the case. Use of a positive ground allows the phone to be driven by the two N-channel transistors inside the CD4001 which are arranged in parallel and are thus able to handle more current for better volume.

INJECTOR/TRACER

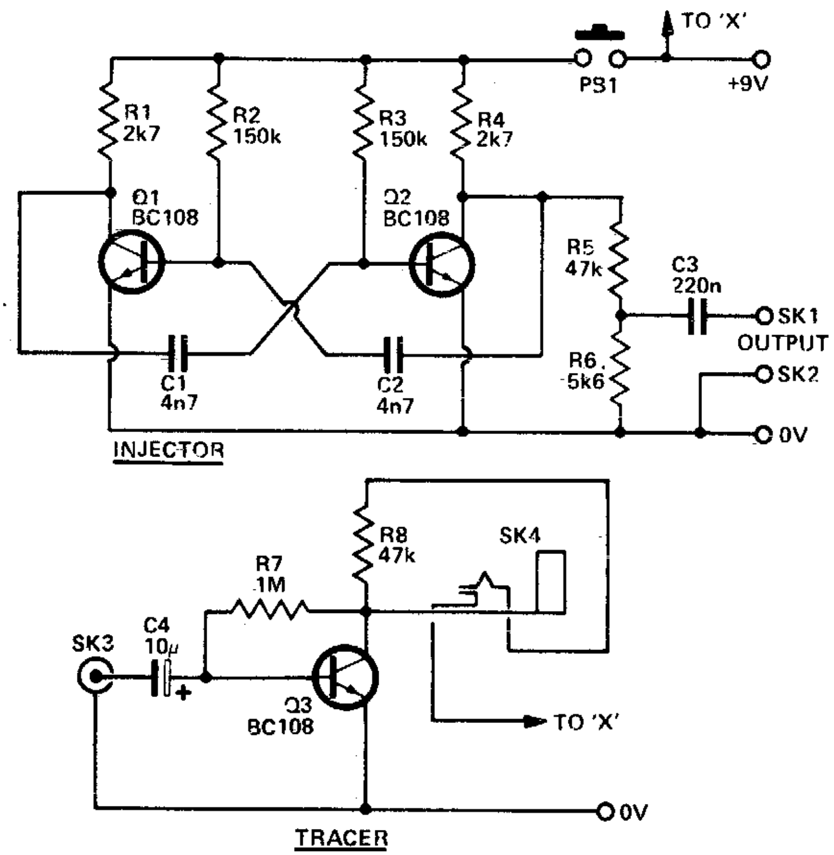


Fig. 67-3

The circuit diagrams for both parts of the injector/tracer. Note that SK4 is used to apply power to the amplifier section.

Circuit Notes

The unit has a separate amplifier and oscillator section allowing them to be used separately if need be. The injector is a multivibrator running at 1 kHz, with R5 and R6 dividing down the output to a suitable level (≈1 V). The tracer is a single-stage amplifier that drives the high impedance earpiece. C4 decouples the input.

CMOS LOGIC PROBE

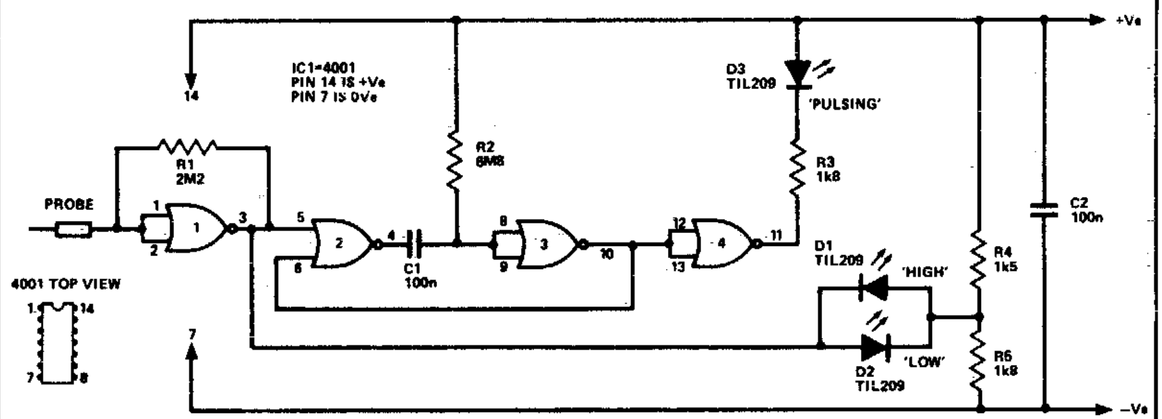


Fig. 67-4

Circuit Notes

The logic probe can indicate four input states, as follows: floating input—all LEDs off; logic 0 input—D2 switched on (D3 will briefly flash on); logic 1 input—D1 switched on; pulsing input—D3 switched on, or pulsing in the case of a low frequency input signal (one or both of the other indicators will switch on, showing if one input state predominates).

RF PROBE FOR VOM

PARTS LIST FOR RF PROBE FOR VOM

C1—500-pF, 400-VDC capacitor
C2—0.001-uF, disc capacitor
D1—1N4149 diode
R1—15,000-ohm, ½-watt resistor

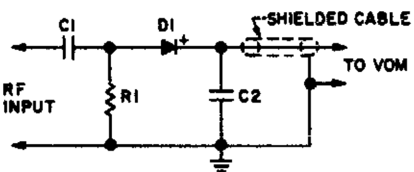


Fig. 67-5

Circuit Notes

This probe makes possible relative measurements of rf voltages to 200 MHz on a 20,000 ohms-per-volt multimeter. Rf voltage must not exceed the breakdown rating of the 1N4149—approximately 100 V.

100 K MEGOHM DC PROBE

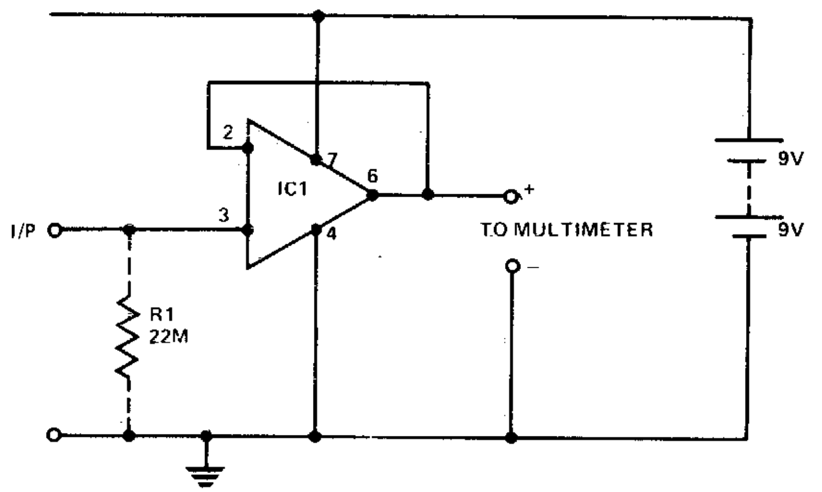


Fig. 67-6

Circuit Notes

A 741 op amp is used with 100% ac and dc feedback to provide a typical input impedance of 10^{11} ohm and unity gain. To avoid hum and rf pickup the input leads should be kept as short as possible and the circuit should be mounted in a small grounded case. Output leads may be long since the output impedance of the circuit is a fraction of an ohm. With no input the output level is indeterminate. Including R1 in the circuit through lowers the input impedance to 22 M.

AUDIBLE TTL PROBE

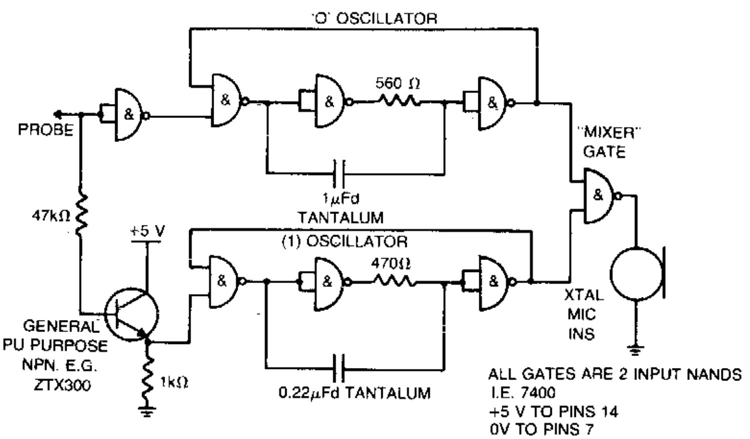


Fig. 67-7

Circuit Notes

When the probe is in contact with a TTL low (0) the probe emits a low note. With a TTL high (1), a high note is emitted. Power is supplied by the circuit under test.

LOGIC PROBE

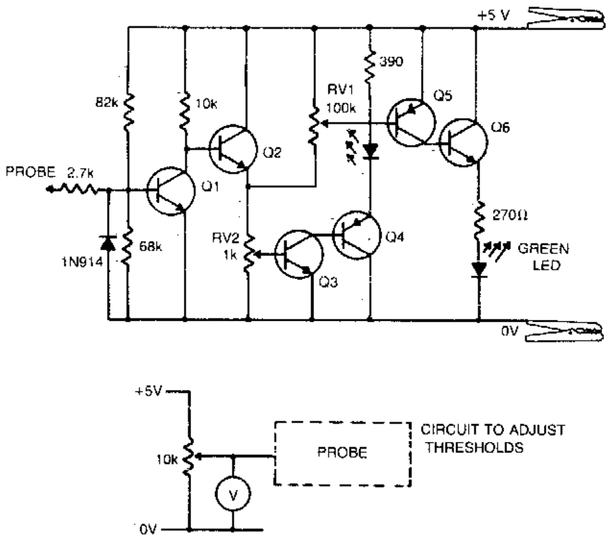


Fig. 67-8

Circuit Notes

Transistors Q1 and Q2 form a buffer, providing the probe with a reasonable input impedance. Q3 and Q4 form a level detecting circuit. As the voltage across the base-emitter junction of the Q3 rises above 0.6 V the transistor turns on thus turning on Q4 and lighting the red (high) LED. Q5 and Q6 perform the same function but for the green (low) LED. Q1, Q4, Q5 are all PNP general purpose silicon transistors (BC178 etc). Q2, Q3, Q6 are all PNP general purpose silicon transistors (BC 108 etc.) The threshold low is ≤ 0.8 V, and the threshold high is ≥ 2.4 V.

LOGIC TEST PROBE WITH MEMORY

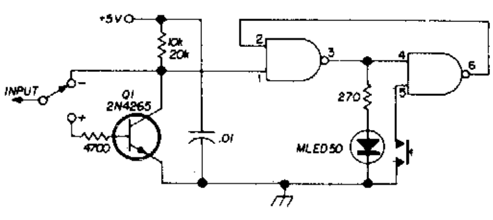


Fig. 67-9

Circuit Notes

There are two switches: a memory disable switch and a pulse polarity switch. Memory disable is a push-button that resets the memory to the low state when depressed. Pulse polarity is a toggle switch that selects whether the probe responds to a high-level or pulse (+5 V) or a low-level or pulse (ground). (Use IC logic of the same type as is being tested).

LOGIC PROBE

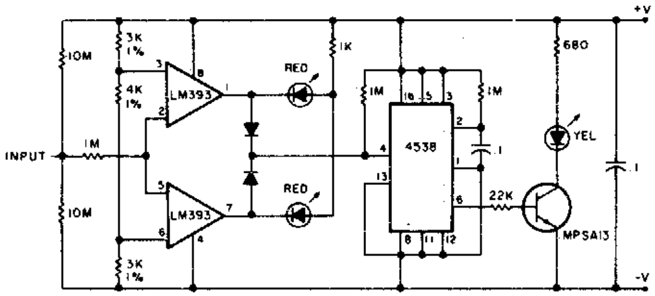


Fig. 67-10

Circuit Notes

The probe indicates a high or low at 70% and 30% of V+ (5 to 12 V). One section of the voltage comparator (LM393) senses V in over 70% of supply and the second section senses V in under 30%. These two sections direct-drive the appropriate LEDs. The pulse detector is a CMOS oneshot (MC14538) triggered on the rising edge of the LM393 outputs through 1N4148 diodes. With the RC values shown, it triggered reliably at greater than 30 kHz on both sine and square waves.

SIMPLE LOGIC PROBE

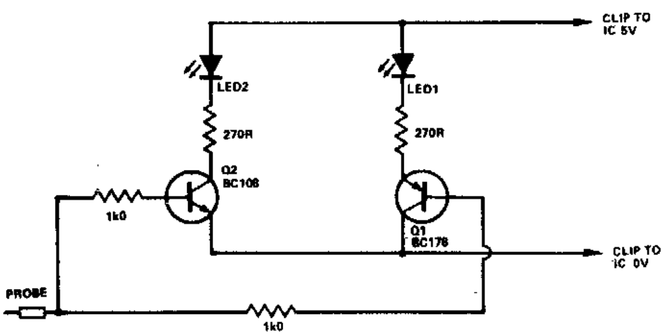
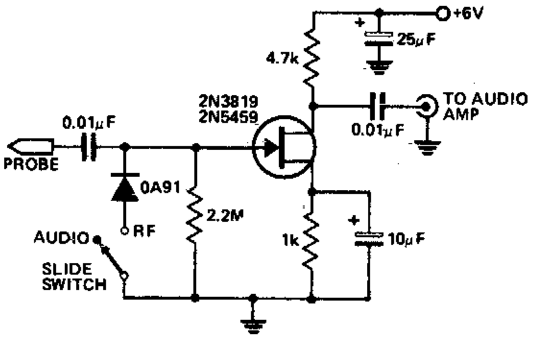


Fig. 67-11

Circuit Notes

If the probe is connected to logic 0, Q1 will be turned on lighting D1. At logic 1, Q2 will be turned on lighting D2. For Q1 and Q2 any NPN or PNP transistors will do. Similarly, D1 and D2 can be any LEDs.

AUDIO-RF SIGNAL TRACER PROBE



Circuit Notes

This economical signal tracer is useful for servicing and alignment work in receivers and low power transmitters. When switched to RF, the modulation on any signal is detected by the diode and amplified by the FET. A twin-core shielded lead can be used to connect the probe to an amplifier and to feed 6 volts to it.

Fig. 67-12

TTL LOGIC TESTER

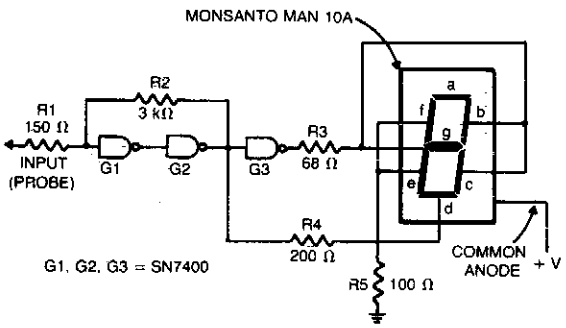


Fig. 67-13

Circuit Notes

Gates G1 and G2 together with resistors R1 and R2 form a simple voltage monitor that has a trip point of 1.4 volts. Gate G3 is simply an inverter. The display section of the tester consists of a common anode alphanumeric LED and current-limiting resistors. It indicates whether the input voltage is above or below 1.4 V, and displays a H or a L (for high or low logic-level) respectively.

68

Pulse Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PULSE GENERATOR

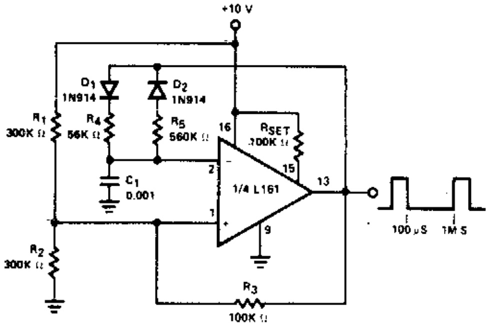


Fig. 68-1

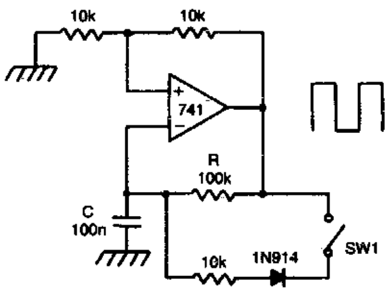
Circuit Notes

The duty cycle of the output pulse is equal to R4/(R4 + R5) × 100%. For duty cycles of less than 50%, D1 can be eliminated and R2 raised according to the following formula:

$$R4(actual) = \frac{R5 \times R4(eff)}{R5 - R4(eff)}$$

R4(eff) is the effective value of R4 in the circuit and R4(actual) is the actual value used; R4(actual) will always be larger than R4(eff).

SINGLE OP AMP OSCILLATOR

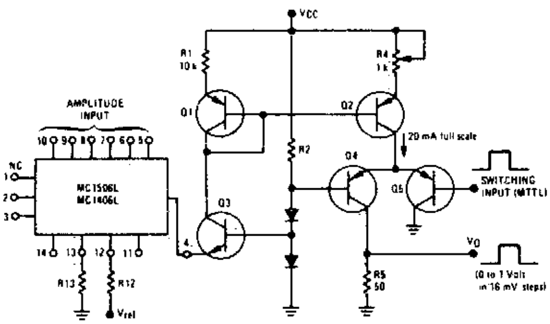


Circuit Notes

This circuit has a Schmitt trigger and integrator built around one op amp. Timing is controlled by the RC network. Voltage at the inverting input follows the RC charging exponential within the upper and lower hysteresis levels. By closing the switch SW1, the discharge time of the capacitor becomes ten times as fast as the rise time. Thus a square wave with an 10:1 mark space ratio is generated.

Fig. 68-2

PROGRAMMABLE PULSE GENERATOR

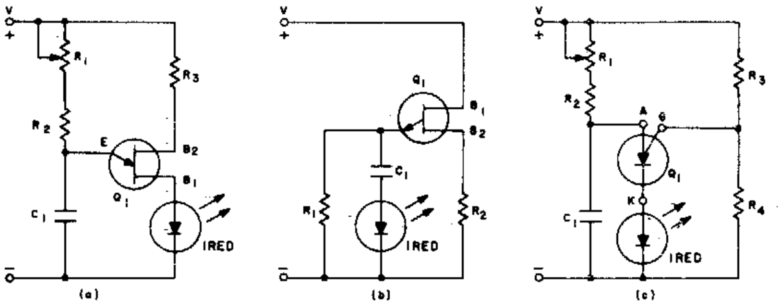


Circuit Notes

Fast rise and fall times require the use of high speed switching transistors for the differential pair, Q4 and Q5. Linear ramps and sine waves may be generated by the appropriate reference input.

Fig. 68-3

UNIJUNCTION TRANSISTOR PULSE GENERATORS



(a) Pulser With Unijunction Transistor (b) Pulser With Complementary Unijunction Transistor (c) Pulser With Programmable Unijunction Transistor

Fig. 68-4

PULSE GENERATOR

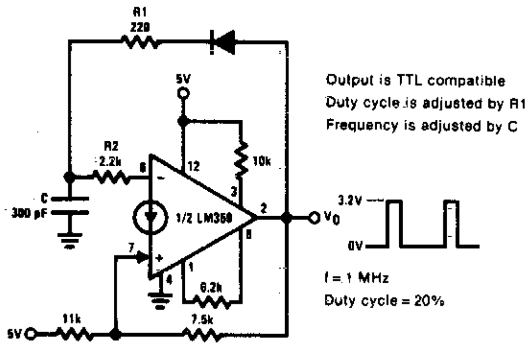


Fig. 68-5

PULSE GENERATOR

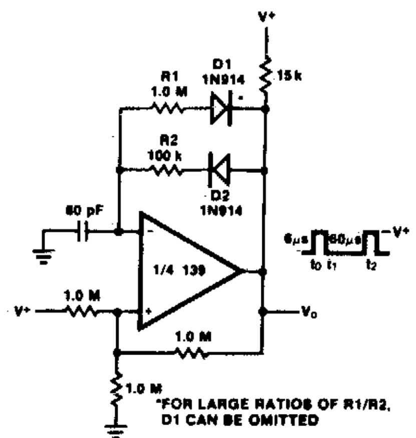


Fig. 68-6

FREE-RUNNING OSCILLATOR

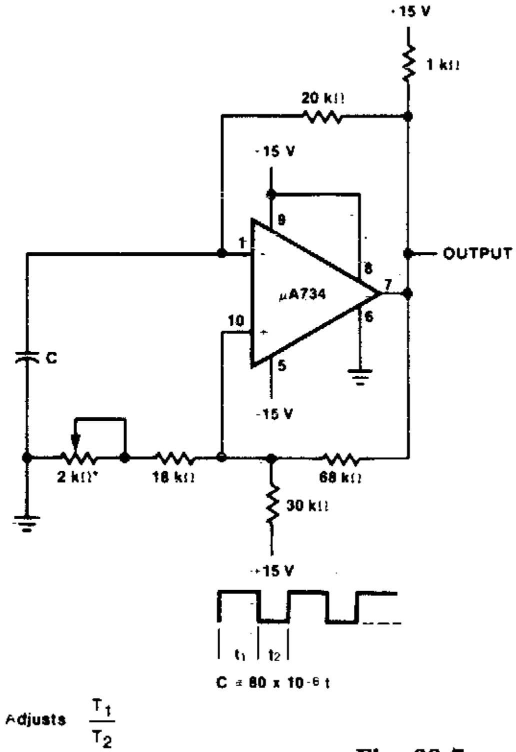


Fig. 68-7

PULSE GENERATOR

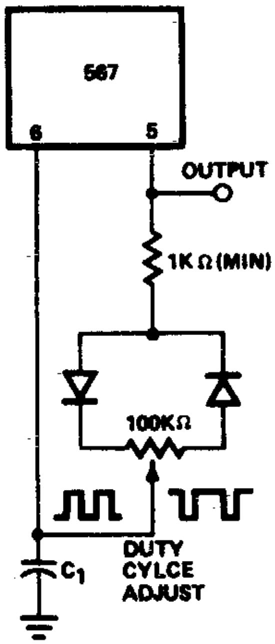


Fig. 68-9

PULSE GENERATOR WITH 25% DUTY CYCLE

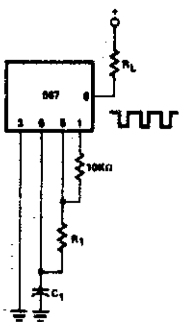
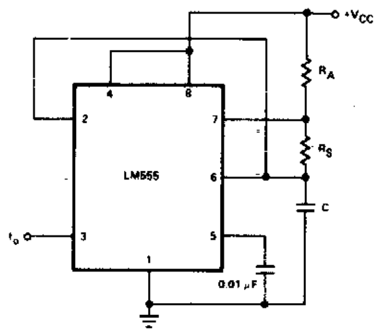


Fig. 68-8

555 TIMER OSCILLATOR

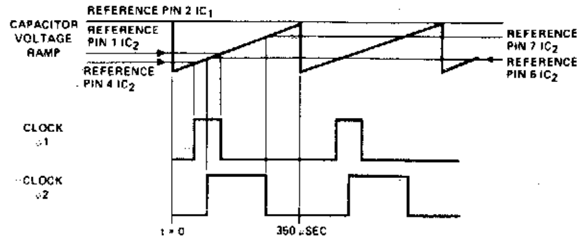


$$f = \frac{1.44}{(R_A + 2\,R_B)\,C}, \quad \text{duty cycle} = \frac{R_B}{R_A + 2\,R_B}$$

a. f = 120 kHz, C = 1200 pF, $R_A = R_B = 10\ k\Omega$

Fig. 68-10

VERSATILE TWO-PHASE PULSE GENERATOR



Typical Waveforms of the Two Phase Pulse Generator

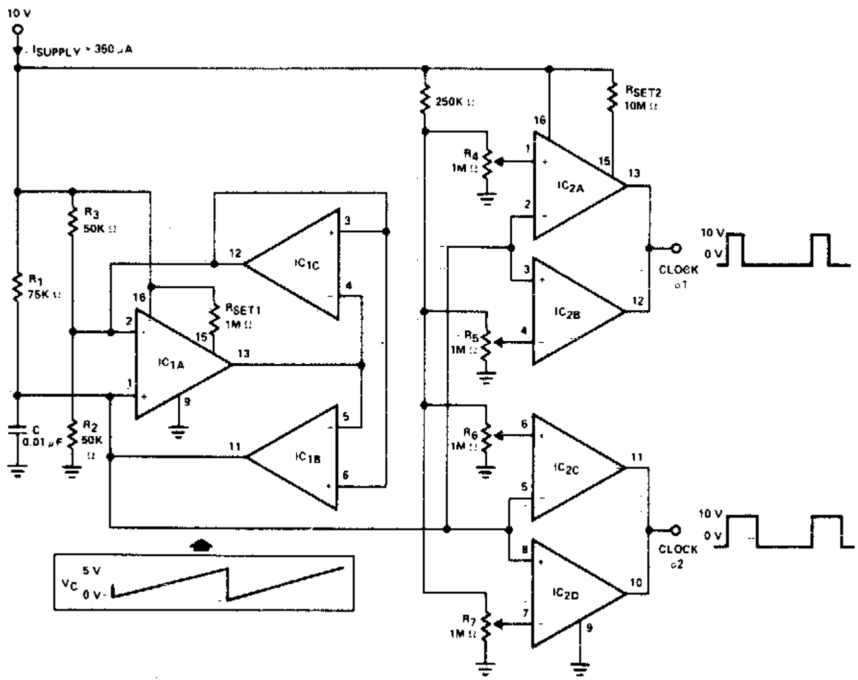


Fig. 68-11

Circuit Notes

Two-phase clock generator uses two L161s to generate pulses of adjustable widths and phase relationships. Ramp generator feeds two variable window comparators formed by IC_{2A}-IC_{2B} and IC_{2C}-IC_{2D} respectively.

69

Radiation Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DOSAGE-RATE METER

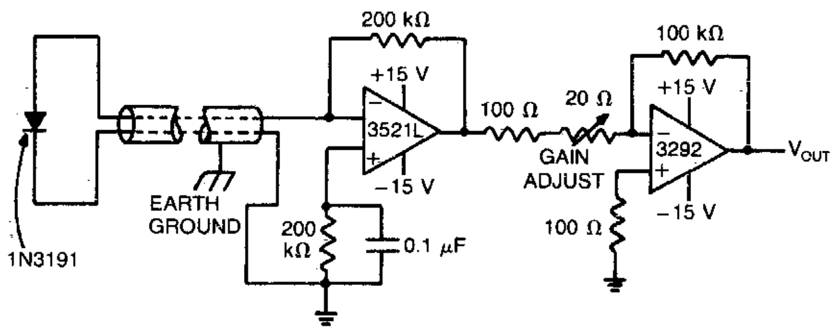
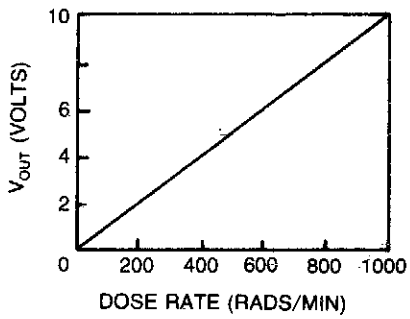


Fig. 69-1



Circuit Notes

A commercial diode is the detector in this highly accurate radiation monitor. The lowdrift FET-input op amp amplifies detector current to a usable level, and the chopper-stabilized amplifier then provides additional gain while minimizing any error caused by ambient-temperature fluctuations. Gain is adjusted so that the output voltage is 1% of incident radiation intensity in rads per minute; therefore voltage can be displayed on 3½ digit DVM for direct reading of dosage rate. Output voltage from the monitor is linearly proportional to radiation intensity at the diode.

WIDEBAND RADIATION MONITOR

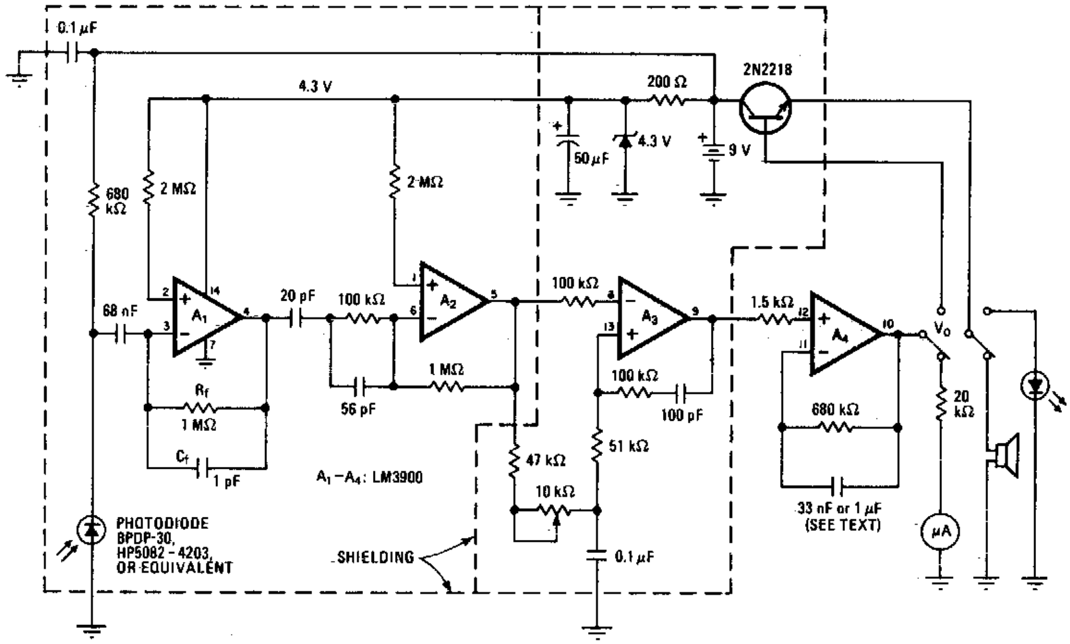


Fig. 69-2

Circuit Notes

A sensitive radiation monitor may be simply constructed with a large-area photodiode and a quad operational amplifier. Replacing the glass window of the diode with Mylar foil will shield it from light and infrared energy, enabling it to respond to such nuclear radiation as alpha and beta particles and gamma rays. A4 integrates the output of A3 in order to drive a microammeter. A 1 microfarad capacitor is used in the integrating network. A lower value, say, 33 nanofarads, will make it possible to drive a small loudspeaker (50-hertz output signal) or light-emitting diode.

GAMMA RAY PULSE INTEGRATOR

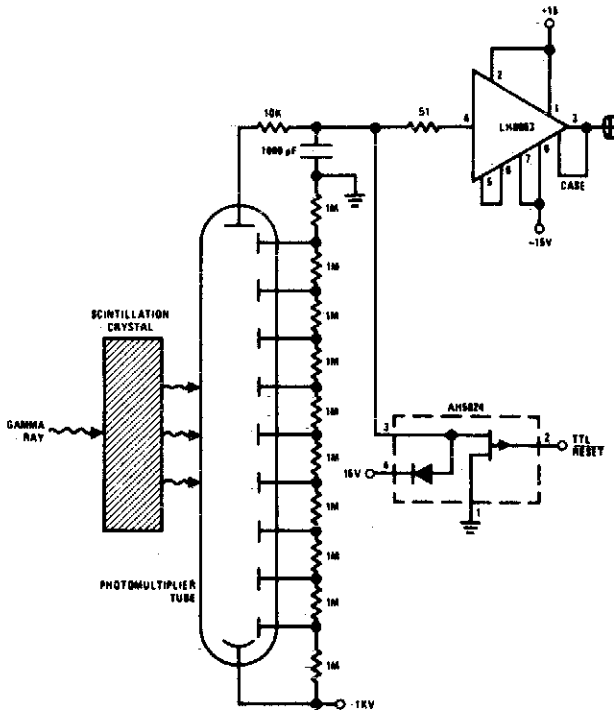


Fig. 69-3

SENSITIVE GEIGER COUNTER

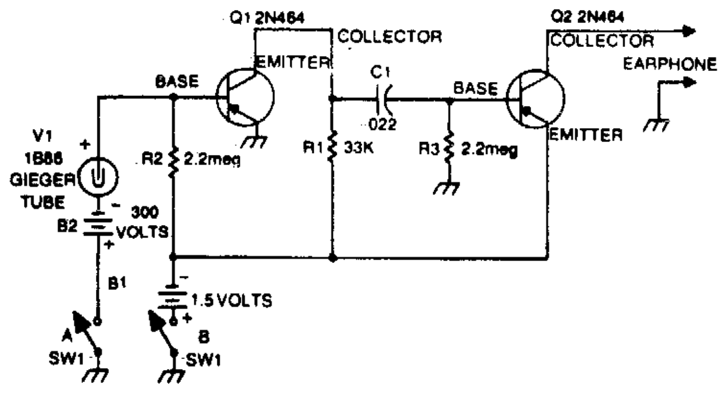


Fig. 69-4

GEIGER COUNTER

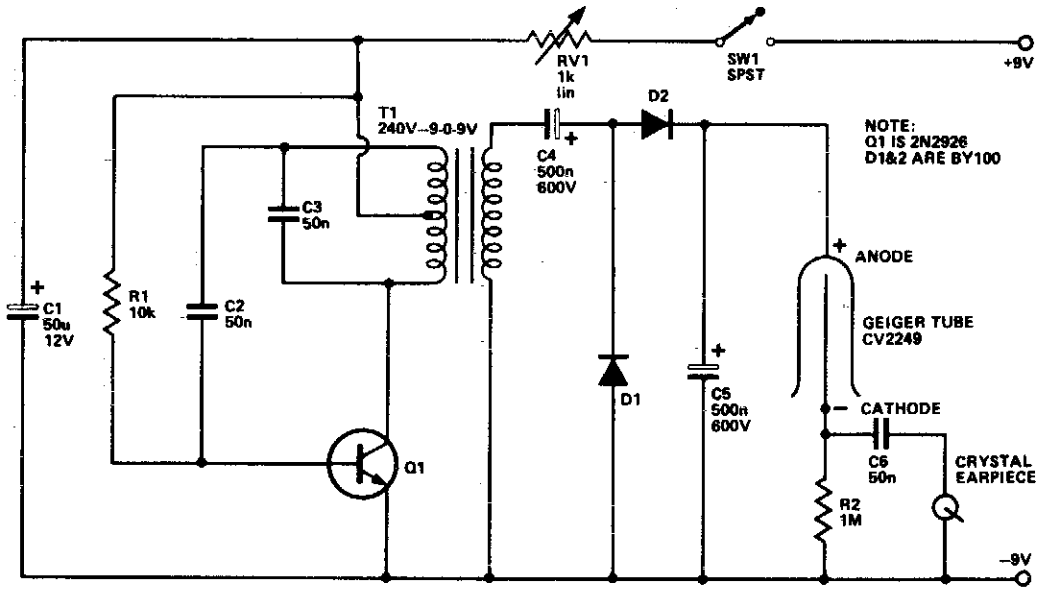


Fig. 69-5

Circuit Notes

The Geiger tube needs a high voltage supply which consists of Q1 and its associated components. The transformer is connected in reverse; the secondary is connected as a Hartley oscillator, and R1 provides base bias. D1, D2, C4, and C5 comprise a voltage doubler. RV1 should be set so that each click heard is nice and clean because over a certain voltage range all that will be heard is a continuous buzz.

NUCLEAR PARTICLE DETECTOR

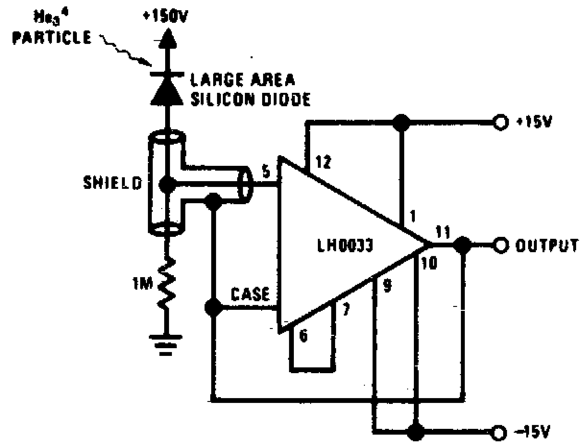


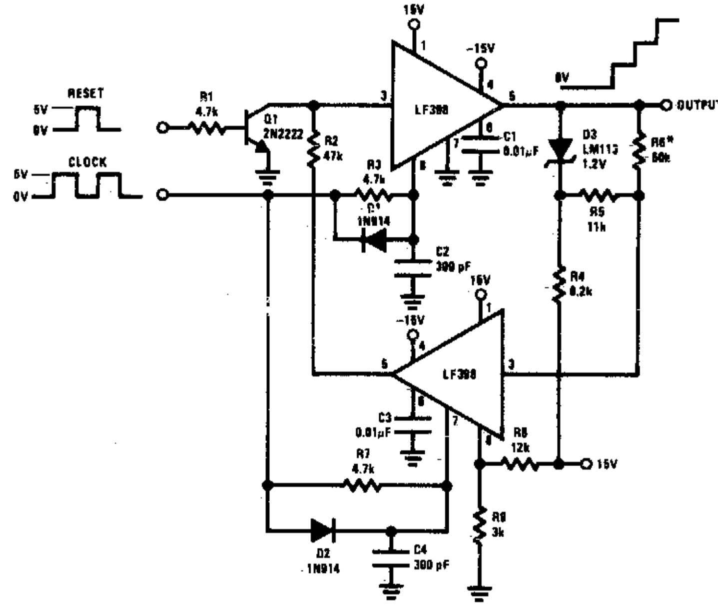
Fig. 69-6

70

Ramp Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

STAIRCASE GENERATOR



*Select for step height
50k → ≈ 1V Step

Fig. 70-1

LINEAR VOLTAGE RAMP GENERATOR

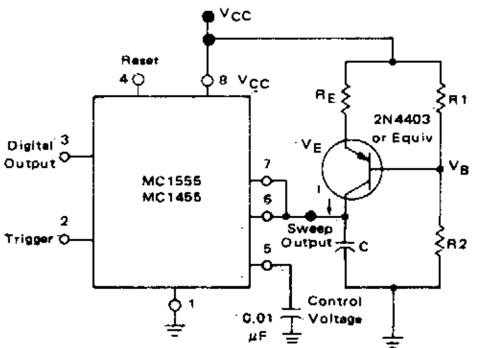


Fig. 70-2

Circuit Notes

In the monostable mode, the resistor can be replaced by a constant current source to provide a linear ramp voltage. The capacitor still charges from 0 to 2/3 Vcc. The linear ramp time is given by the following equation:

$$I = \frac{V_{CC} - V_B - V_{BE}}{R_E} \qquad t = \frac{2}{3}\frac{V_{CC}}{I}$$

If V_B is much larger than V_{BE}, then t can be made independent of V_{CC}.

PRECISION RAMP GENERATOR

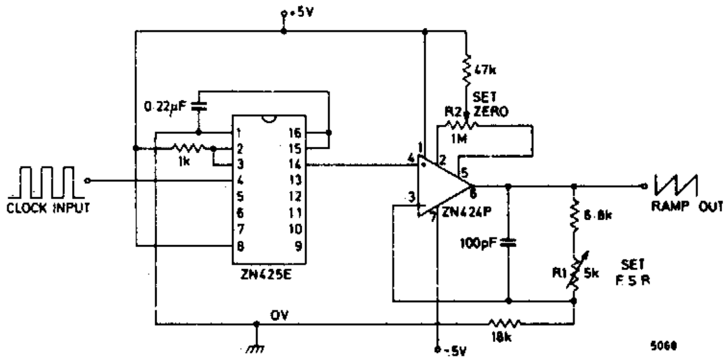
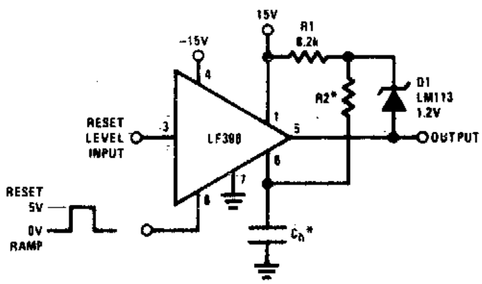


Fig. 70-3

RAMP GENERATOR WITH VARIABLE RESET LEVEL



*Select for ramp rate $\frac{\Delta V}{\Delta T} = \frac{1.2V}{(R2)(C_h)}$
R ≥ 10k

Fig. 70-4

71

Receivers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Clock Radio
AM/FM Clock Radio
AM Radio
FM Stereo Demodulation System
Analog Receiver
FM Radio
Simple LF Converter
CMOS Line Receiver
Squelch Circuit for AM or FM
VLF Converter

CLOCK RADIO

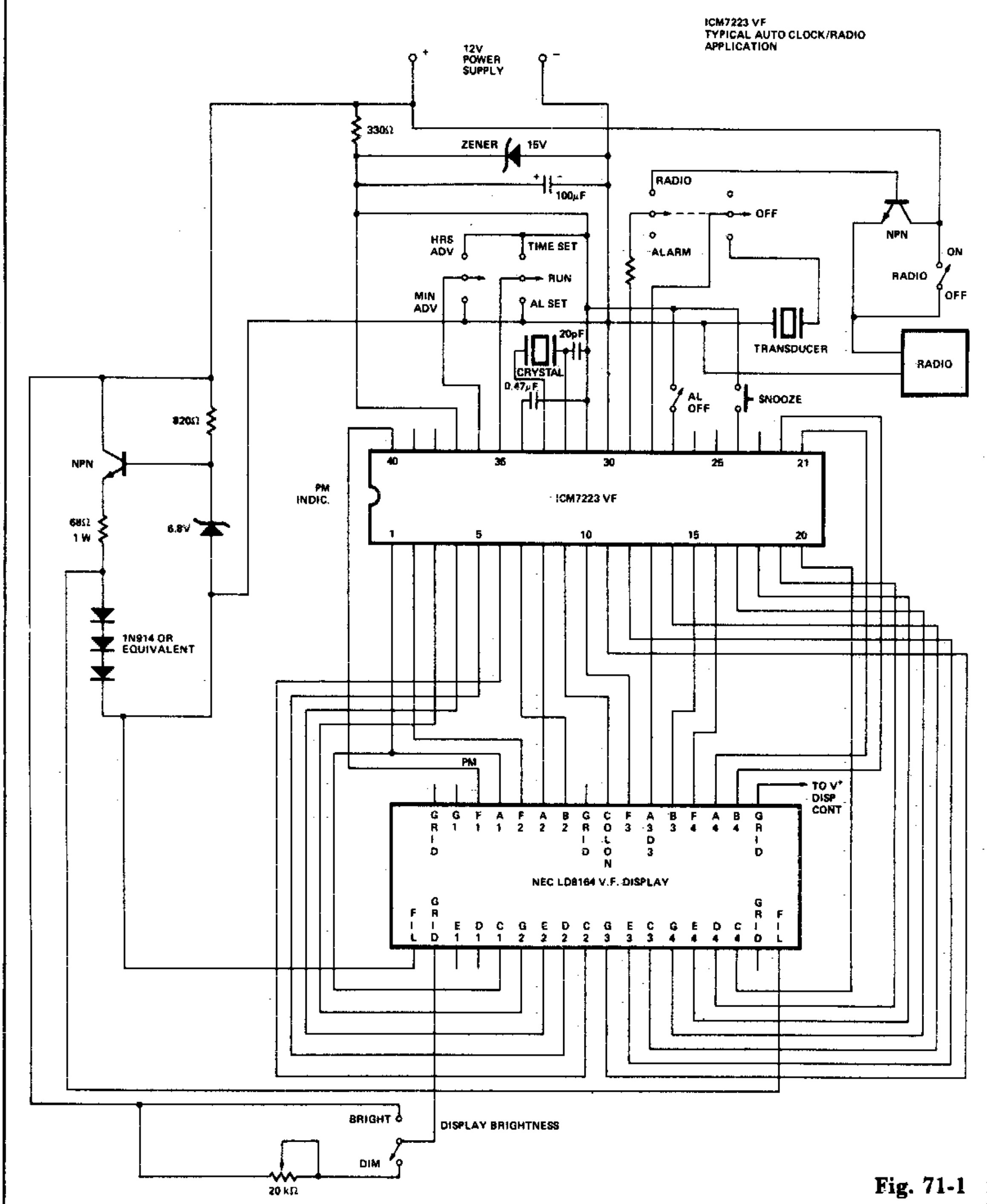
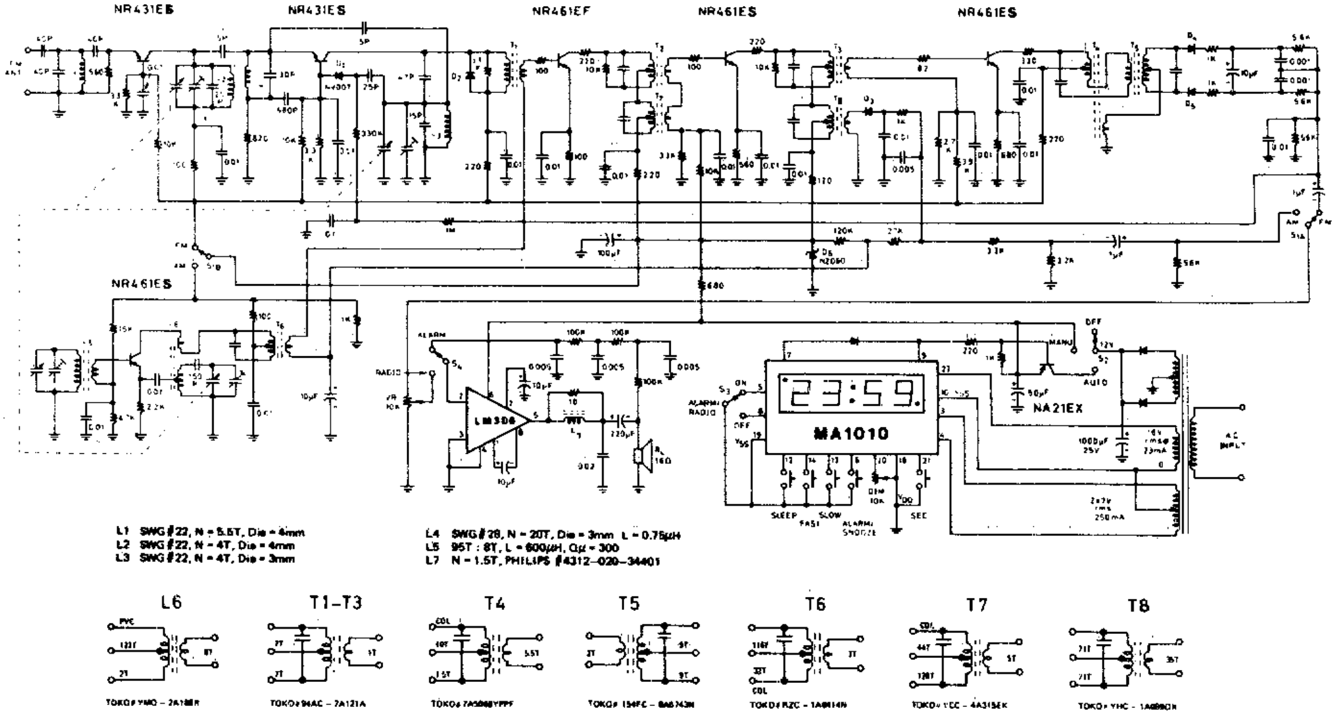


Fig. 71-1

AM/FM CLOCK RADIO



FM performance (88–108 MHz)

- 30dB quieting sensitivity: 5μV
- limiting sensitivity: 20μV
- AM rejection: 40dB
- AFC holding range: 800KHz
- Bandwidth: 180KHz

AM performance (525–1650 KHz)

- maximum sensitivity: 100μV/M
- 20dB quieting sensitivity: 280μV/M
- selectivity ± 10KHz: –28dB
- AGC figure of merit: 40dB
- overload distortion: 6%

AUDIO performance

- gain at 1 KHz: 200
- 10% THD output power: 900mW
- frequency response: 70Hz – 12KHz
- typical system dist: 0.8%
- alarm tone frequency: 600Hz

Fig. 71-2

AM RADIO

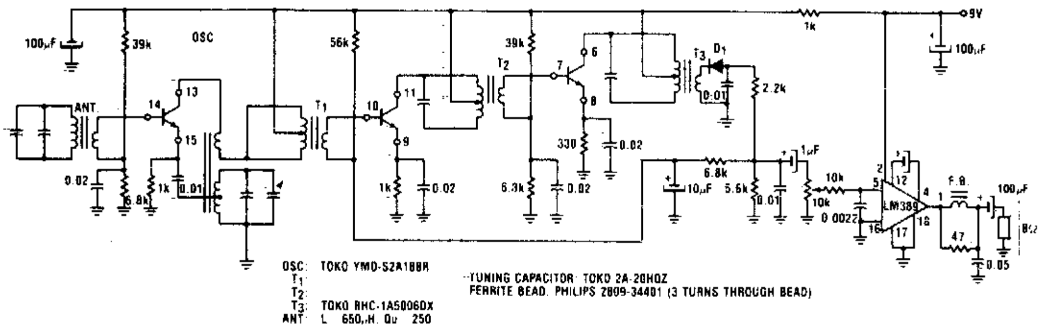


Fig. 71-3

FM STEREO DEMODULATION SYSTEM

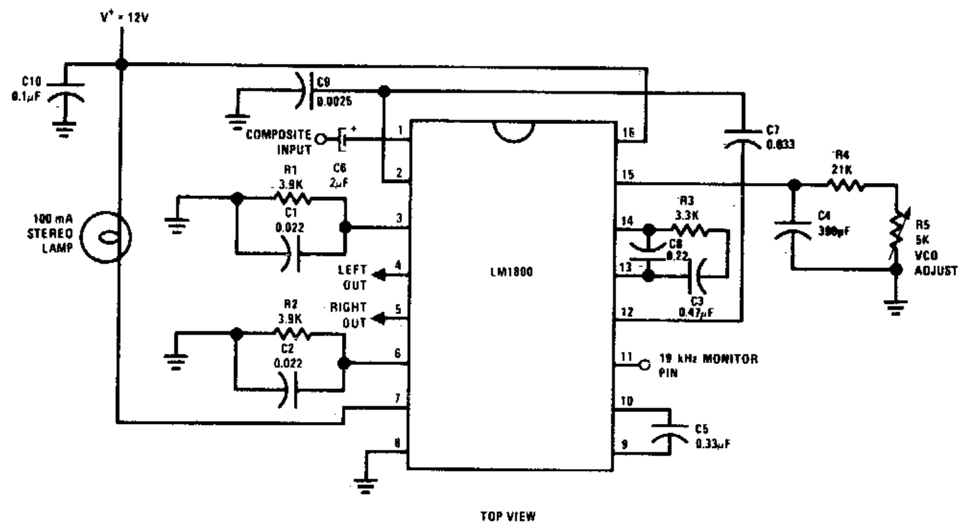


Fig. 71-4

ANALOG RECEIVER (LOW TEMPERATURE DRIFT)

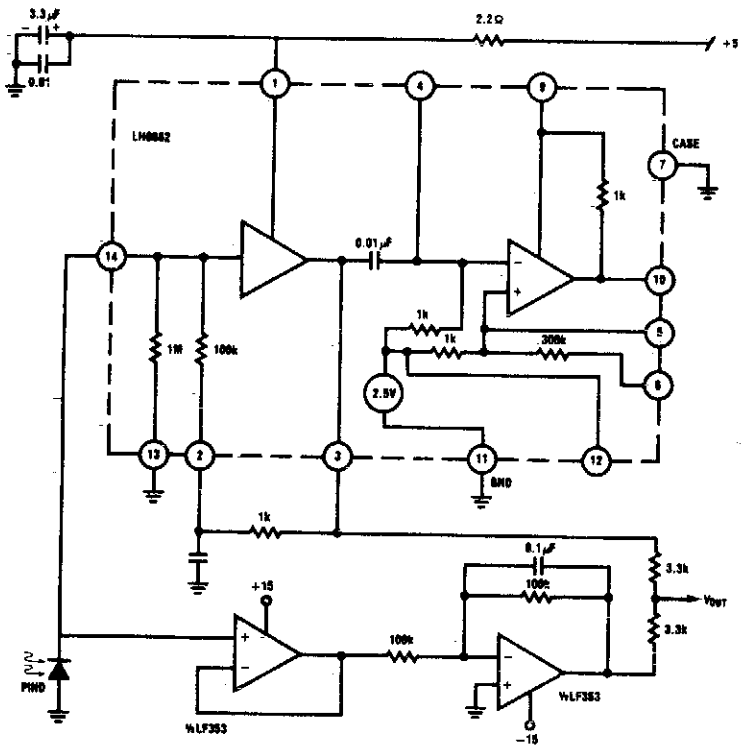


Fig. 71-5

FM RADIO

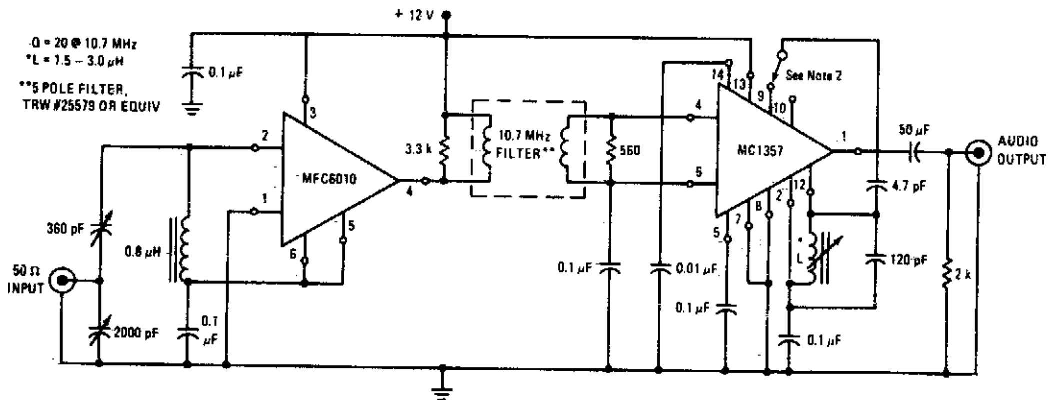


Fig. 71-6

SIMPLE LF CONVERTER

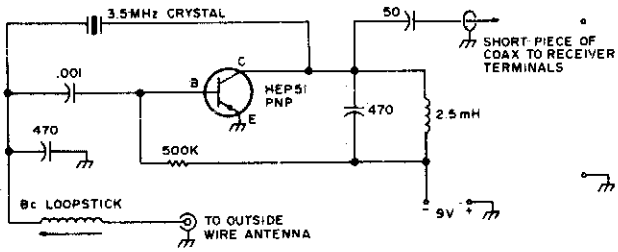
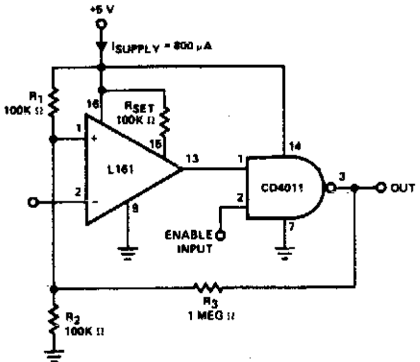


Fig. 71-7

Circuit Notes

This converter allows coverage from 25 kHz up to 500 kHz. Use short coax from the converter to receiver antenna input. Tune the receiver to 3.5 MHz, peak for loudest crystal calibrator and tune your receiver higher in frequency to 3.6 MHz and you're tuning the 100 kHz range. 3.7 MHz puts you at 200 kHz, 3.8 MHz equals 300 kHz, 3.9 MHz yields 500 kHz, and 4.0 MHz gives you 500 kHz.

CMOS LINE RECEIVER



Circuit Notes

The trip point is set half way between the supplies by R1 and R2; R3 provides over 200 mV of hysteresis to increase noise immunity. Maximum frequency of operation is about 300 kHz. If response to TTL levels is desired, change R2 to 39 K. The trip point is now centered at 1.4 V.

Fig. 71-8

SQUELCH CIRCUIT FOR AM OR FM

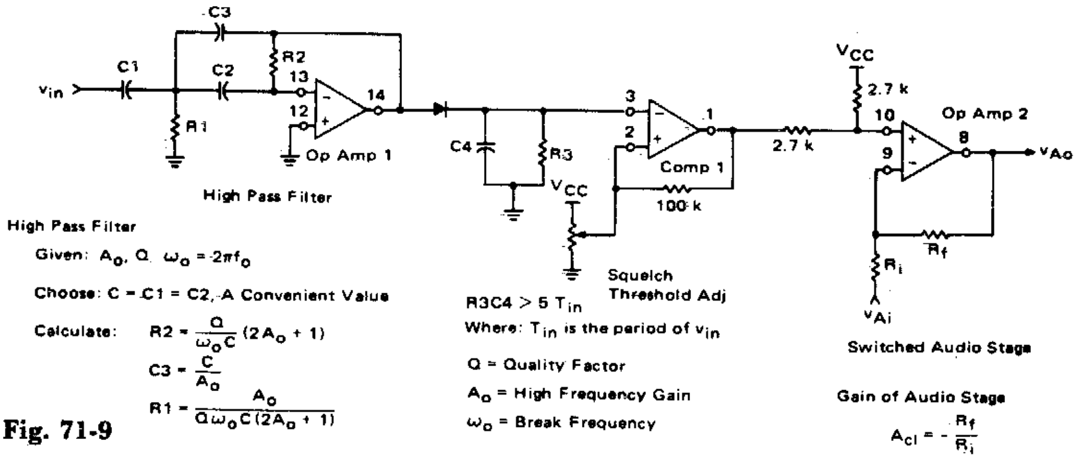


Fig. 71-9

VLF CONVERTER

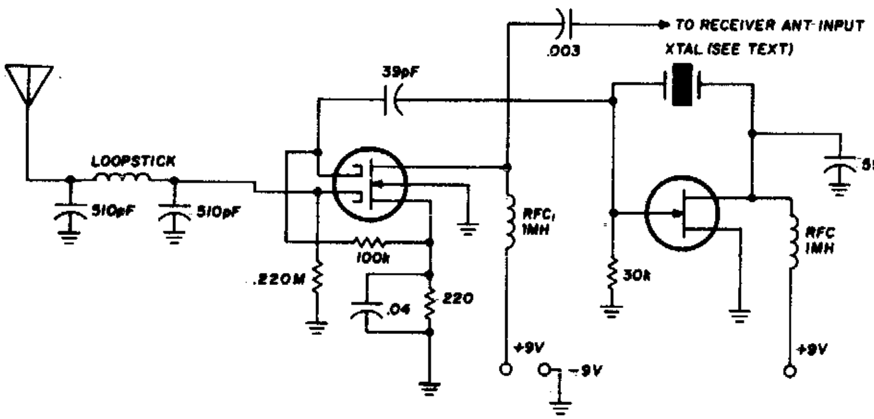


Fig. 71-10

Circuit Notes

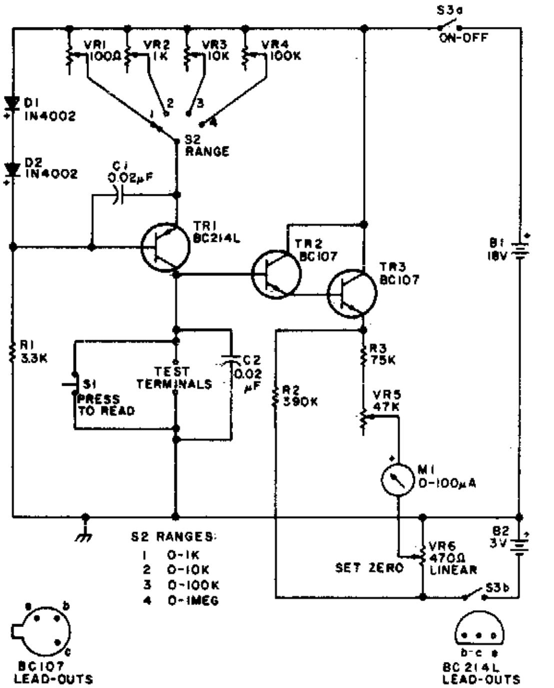
This converter uses a low-pass filter instead of the usual tuned circuit so the only tuning required is with the receiver. The dual-gate MOSFET and FET used in the mixer and oscillator aren't critical. Any crystal having a frequency compatible with the receiver tuning range may be used. For example, with a 3500 kHz crystal, 3500 kHz on the receiver dial corresponds to zero kHz; 3600 to 100 kHz; 3700 to 200 kHz, etc. (At 3500 khz on the receiver all one can hear is the converter oscillator, and VLF signals start to come in about 20 kHz higher.)

72

Resistance and Continuity Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LINEAR SCALE OHMMETER



Circuit Notes

This circuit is designed to provide accurate measurement and a linear resistance scale at the high end. The circuit has four ranges. Another meter with a current range of 10 μA to 10 mA and sensitivity of 10,000 ohms per volt is needed for setting up.

Fig. 72-1

OHMMETER

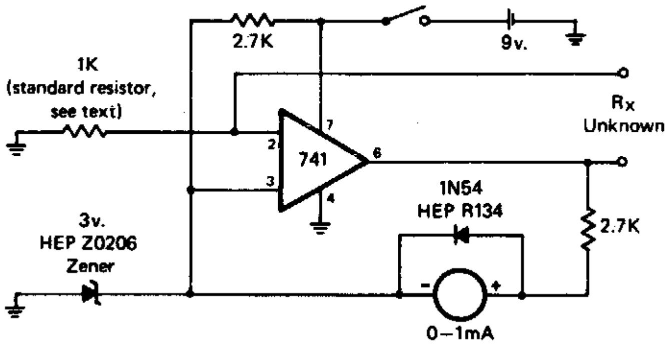


Fig. 72-2

Circuit Notes

This circuit has a linear reading scale, requires no calibration, and requires no zero adjustment. It may be made multirange by switching in different standard resistors.

LOW PARTS COUNT RATIOMETRIC RESISTANCE MEASUREMENT

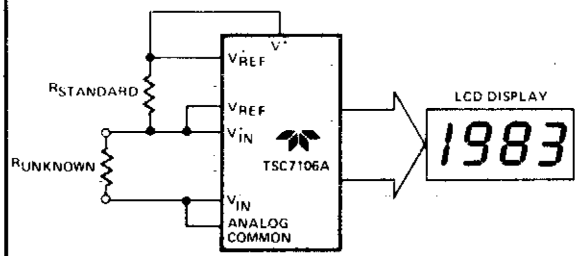


Fig. 72-3

Circuit Notes

The unknown resistance is put in series with a known standard and a current passed through the pair. The voltage developed across the unknown is applied to the input and the voltage across the known resistor applied to the reference input. If the unknown equals the standard, the display will read 1000. The displayed reading can be determined from the following expression:

$$\text{Displayed Reading} = \frac{R_{unknown}}{R_{standard}} \times 1000$$

The display will overrange for $R_{unknown}$, $\geq 2 \times R_{standard}$.

AUDIO CONTINUITY TESTER

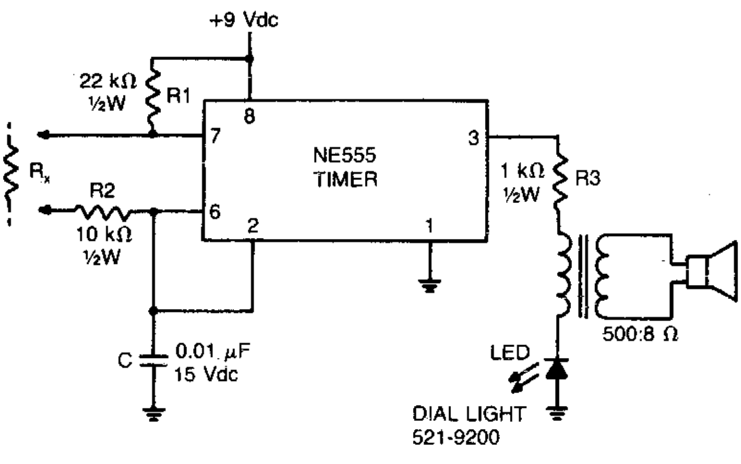


Fig. 72-4

Circuit Notes

This low-current audio continuity tester indicates the unknown resistance value by the frequency of audio tone. A high tone indicates a low resistance, and a tone of a few pulses per second indicates a resistance as high as 30 megohms.

LOW RESISTANCE CONTINUITY TESTER

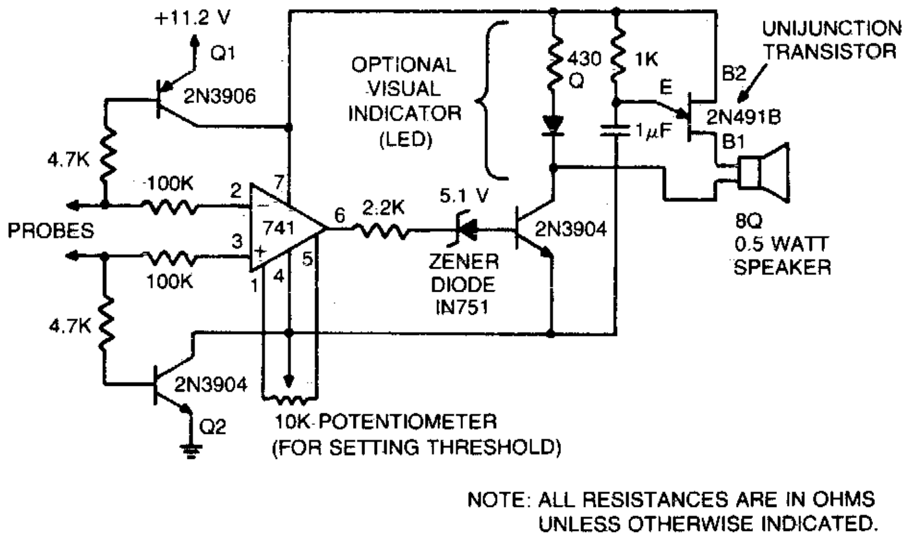


Fig. 72-5

Circuit Notes

This tester can be used to check IC printed circuit boards. Two 4.7 K resistors and the transistors connected to them prevent current flow through the operational amplifier until the probe circuit is completed. The zener diode in series with the operational amplifier output prevents audio oscillator operation until the positive output of the operational amplifier has sufficient amplitude.

"BUZZ BOX" CONTINUITY AND COIL CHECKER

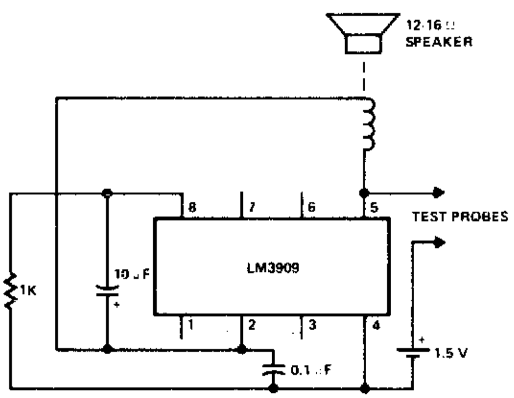


Fig. 72-6

Circuit Notes

Differences between shorts, coils, and a few ohms of resitance can be heard.

LINEAR SCALE OHMMETER

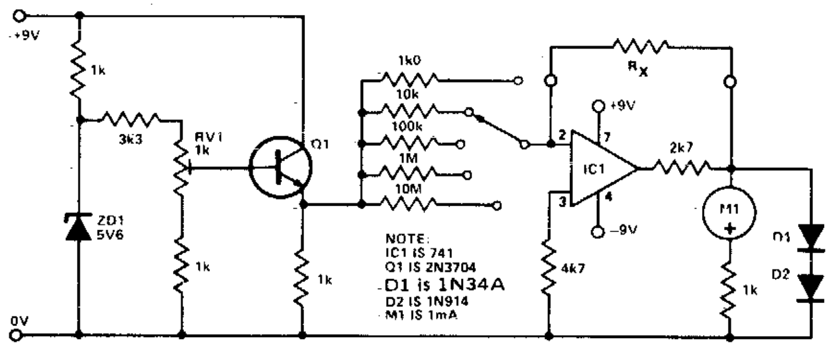


Fig. 72-7

Circuit Notes

One preset resistor is used for all the ranges, simplifying the setting up. Diode clamping is included to prevent damage to the meter if the unknown resistor is higher than the range selected. When the meter has been assembled, a 10 K precision resistor is placed in the test position, R_x; the meter is set to the 10 K range and RV1 is adjusted for full scale deflection.

BRIDGE CIRCUIT

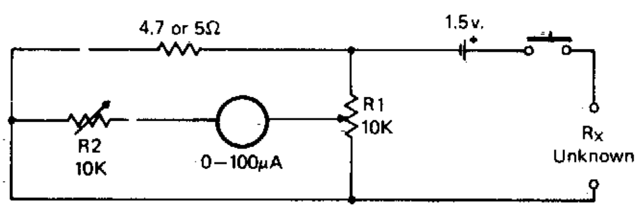


Fig. 72-8

Circuit Notes

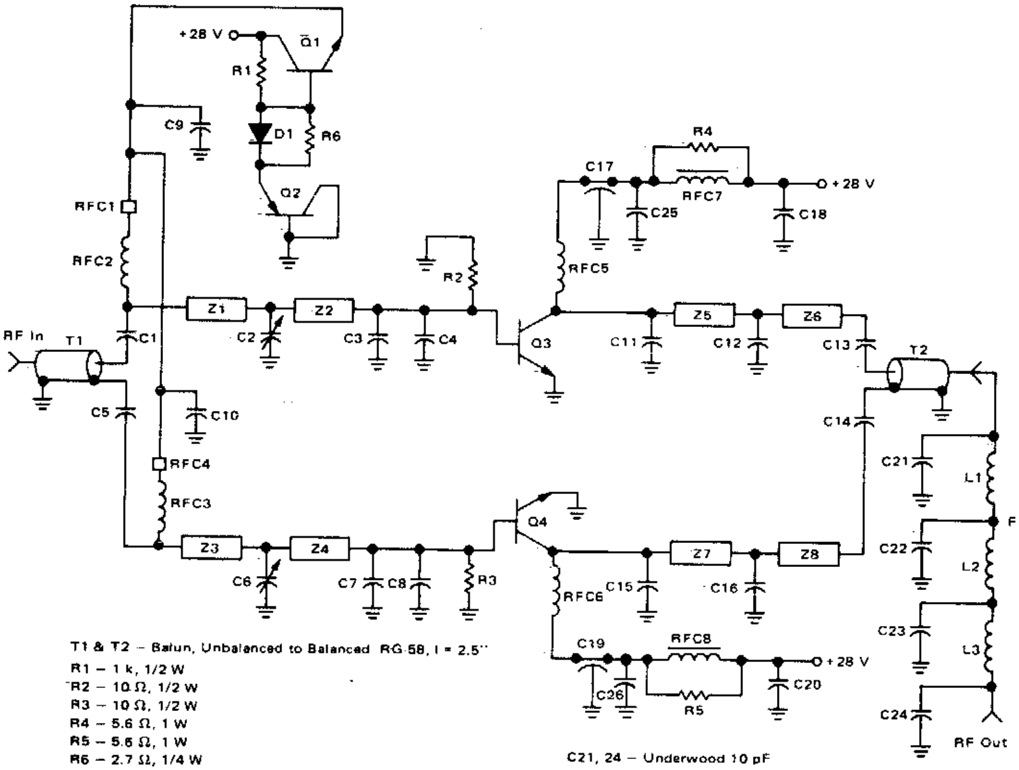
For measurement of resistances from about 5 ohms down to about 1/10 ohm.

73

RF Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

100 W PEP 420-450 MHz PUSH-PULL LINEAR AMPLIFIER



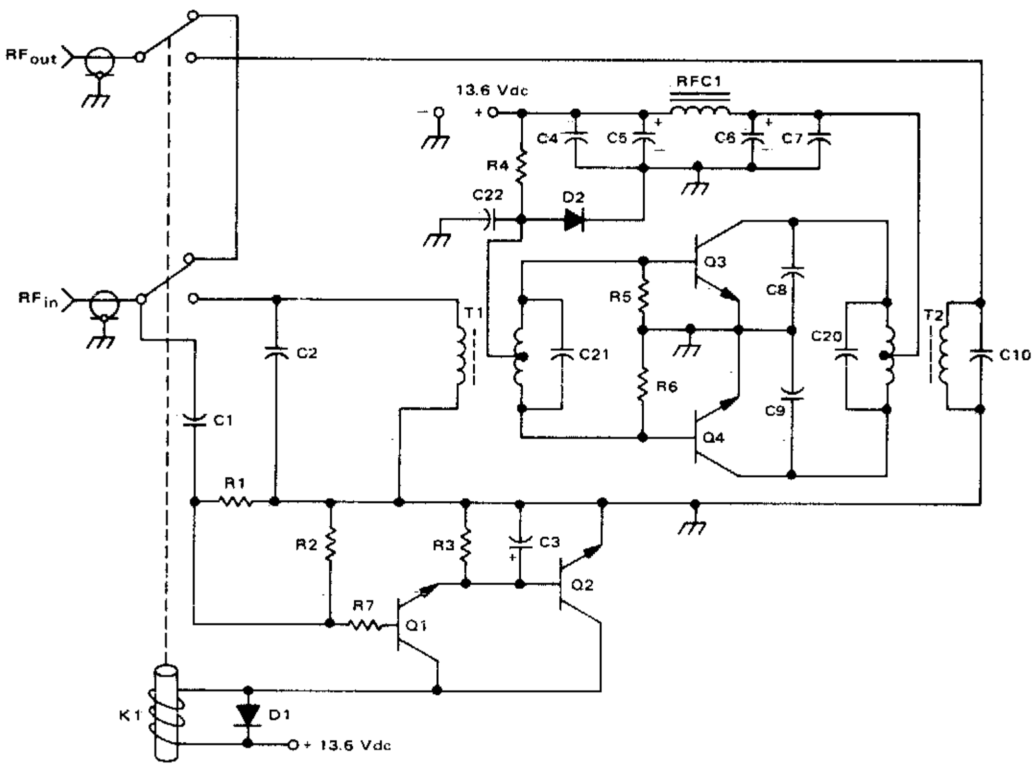
T1 & T2 – Balun, Unbalanced to Balanced RG-58, l = 2.5"
R1 – 1 k, 1/2 W
R2 – 10 Ω, 1/2 W
R3 – 10 Ω, 1/2 W
R4 – 5.6 Ω, 1 W
R5 – 5.6 Ω, 1 W
R6 – 2.7 Ω, 1/4 W
Z1 & Z3 – Microstrip – W = 200 mils, l = 1.8"
Z2 & Z4 – Microstrip – W = 200 mils, l = 300 mils
Z5 & Z6 – Microstrip – W = 150 mils, l = 300 mils
Z6 & Z8 – Microstrip – W = 150 mils, l = 1.4"
RFC1, 4 – Ferroxcube Bead 56-590-65-3B
RFC2,3 – 0.15 μH Cambion Molded Coil
RFC5, 6 – 1 Turn #20 Enameled Wire Wound on 5/16" Bolt
RFC7, 8 – VK200 20/4B
C1,3,4,5,7,8,11,15 – Underwood 40 pF
C12,16 – Underwood 25 pF
C13, 14, 22, 23 – Underwood 15 pF
C9,10,18,20 – 1 μF Tantalum
C21, 24 – Underwood 10 pF
C2,6 – Arco 403
C17,19 – Underwood J102, 1000 pF Feed Thru
C25, 26 – 0.1 μF, Erie Red Cap
L1 – 24 nH, #14 Wire, l = 1.2"
L2 – 12 nH, #14 Wire, l = 0.6"
L3 – 24 nH, #14 Wire, l = 1.2"
Board – G10, ϵR ≈ 5, t = 0.062", l = 8.0", W = 4.0"
Q1 – 2N5192
Q2 – 2N5194
Q3 – MRF309
Q4 – MRF309
D1 – 1N4001

Fig. 73-1

Circuit Notes

This 100 watt linear amplifier may be constructed using two MRF309 transistors in push-pull, requiring only 16 watts drive from 420 to 450 MHz. Operating from a 28 volt supply, eight dB of power gain is achieved along with excellent practical performance featuring: maximum input SWR of 2:1, harmonic suppression more than—63 dB below 100 watts output, efficiency greater than 40%, circuit stability with a 3:1 collector mismatch at all phase angles.

140 W (PEP) AMATEUR RADIO LINEAR AMPLIFIER (2-30 MHz)



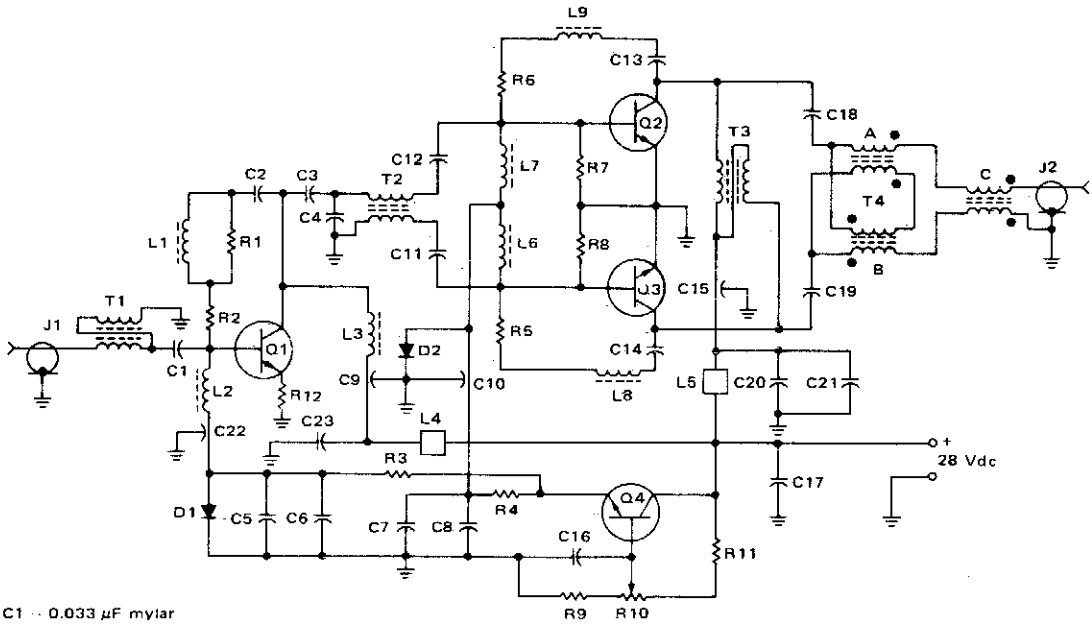
C1	=	33 pF Dipped Mica
C2	=	18 pF Dipped Mica
C3	=	10 µF 35 Vdc for AM operation, 100 µF 35 Vdc for SSB operation.
C4	=	.1 µF Erie
C5	=	10 µF 35 Vdc Electrolytic
C6	=	1 µF Tantalum
C7	=	.001 µF Erie Disc
C8, 9	=	330 pF Dipped Mica
R1	=	100 kΩ 1/4 W Resistor
R2, 3	=	10 kΩ 1/4 W Resistor
R4	=	33 Ω 5 W Wire Wound Resistor
R5, 6	=	10 Ω 1/2 W Resistor
R7	=	100 Ω 1/4 W Resistor
RFC1	=	9 Ferroxcube Beads on #18 AWG Wire
D1	=	1N4001
D2	=	1N4997
Q1, Q2	=	2N4401
Q3, 4	=	MRF454
T1, T2	=	16:1 Transformers
C20	=	910 pF Dipped Mica
C21	=	1100 pF Dipped Mica
C10	=	24 pF Dipped Mica
C22	=	500 µF 3 Vdc Electrolytic
K1	=	Potter & Brumfield KT11A 12 Vdc Relay or Equivalent

Fig. 73-2

Circuit Notes

This inexpensive, easy to construct amplifier uses two MRF454 devices. Specified at 80 W power output with 5 W of input drive, 30 MHz, and 12.5 Vdc.

160 W (PEP) BROADBAND LINEAR AMPLIFIER



C1 -- 0.033 μF mylar
C2, C3 – 0.01 μF mylar
C4 – 620 pF dipped mica
C5, C7, C16 – 0.1 μF ceramic
C6 – 100 μF/15 V electrolytic
C8 -- 500 μF/6 V electrolytic
C9, C10, C15, C22 -- 1000 pF feed through
C11, C12 – 0.01 μF
C13, C14 – 0.015 μF mylar
C17 – 10 μF/35 V electrolytic
C18, C19, C21 -- Two 0.068 μF mylars in parallel
C20 – 0.1 μF disc ceramic
C23 - 0.1 μF disc ceramic
R1 – 220 Ω, 1/4 W carbon
R2 – 47 Ω, 1/2 W carbon
R3 – 820 Ω, 1 W wire W
R4 – 35 Ω, 5 W wire W
R5, R6 – Two 150 Ω, 1/2 W carbon in parallel
R7, R8 – 10 Ω, 1/2 W carbon
R9, R11 – 1 k, 1/2 W carbon
R10 – 1 k, 1/2 W potentiometer
R12 – 0.85 Ω (6 5.1 Ω or 4 3.3 Ω 1/4 W resistors in parallel, divided equally between both emitter leads)

T1 – 4:1 Transformer, 6 turns, 2 twisted pairs of #26 AWG enameled wire (8 twists per inch)

T2 -- 1:1 Balun, 6 turns, 2 twisted pairs of #24 AWG enameled wire (6 twists per inch)

T3 – Collector choke, 4 turns, 2 twisted pairs of #22 AWG enameled wire (6 twists per inch)

T4 – 1:4 Transformer Balun, A&B – 5 turns, 2 twisted pairs of #24, C – 8 turns, 1 twisted pair of #24 AWG enameled wire (All windings 6 twists per inch). (T4 – Indiana General F624-19Q1, – All others are Indiana General F627-8Q1 ferrite toroids or equivalent.)

PARTS LIST

L1 – .33 μH, molded choke
L2, L6, L7 – 10 μH, molded choke
L3 – 1.8 μH (Ohmite 2-144)
L4, L5 – 3 ferrite beads each
L8, L9 – .22 μH, molded choke

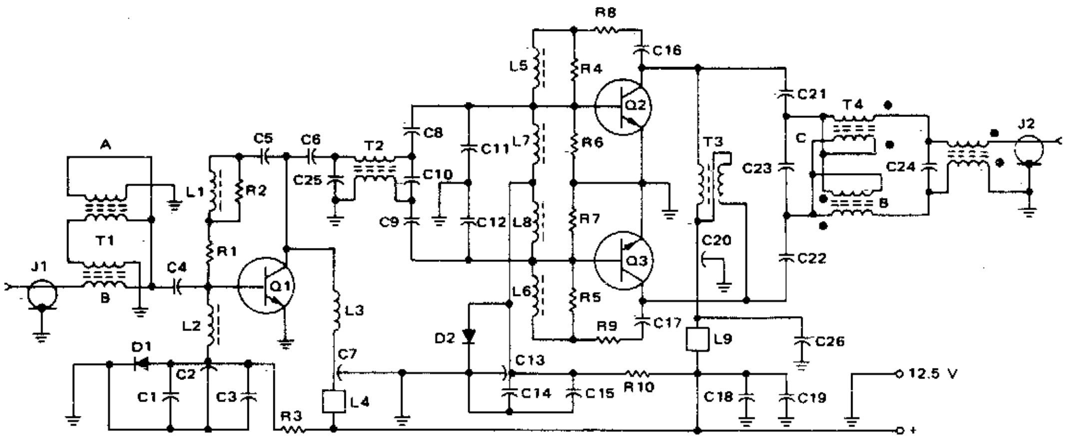
Q1 -- 2N6370
Q2, Q3 – 2N5942
Q4 – 2N5190

D1 – 1N4001
D2 – 1N4997

J1, J2 – BNC connectors

Fig. 73-3

80 W (PEP) BROADBAND/LINEAR AMPLIFIER



C1, C14, C18 – 0.1 μF ceramic.
C2, C7, C13, C20 – 0.001 μF feed through.
C3 – 100 μF/3V.
C4, C6 – 0.033 μF mylar
C5 – 0.0047 μF mylar.
C8, C9 – 0.015 and 0.033 μF mylars in parallel.
C10 – 470 pF mica.
C11, C12 – 560 pF mica.
C15 – 1000 μF/3 V
C16, C17 – 0.015 μF mylar
C19 – 10 pF 15 V
C21, C22 – two 0.068 μF mylars in parallel.
C23 – 330 pF mica
C24 – 39 pF mica
C25 – 680 pF mica
C26 – .01 μF ceramic

R1, R6, R7 – 10 Ω, 1/2 W carbon.
R2 – 51 Ω, 1/2 W carbon
R3 – 240 Ω, 1 wire W
R4, R5 – 18 Ω, 1 W carbon
R8, R9 – 27 Ω, 2 W carbon
R10 – 33 Ω, 6 W wire W

L1 – 0.22 μh molded choke
L2, L7, L8 – 10 μh molded choke
L5, L6 – 0.15 μh
L3 – 25 t, #26 wire, wound on a 100 Ω, 2 W resistor. (1.0 μh)
L4, L9 – 3 ferrite beads each.

T1 – 2 twisted pairs of #26 wire, 8 twists per inch. A = 4 turns, B = 8 turns. Core- -Stackpole 57-9322-11, Indiana General F627-8Q1 or equivalent

T2 – 2 twisted pairs of #24 wire, 8 twists per inch, 6 turns. (Core as above.)

T3 – 2 twisted pairs of #20 wire, 6 twists per inch, 4 turns. (Core as above.)

T4 – A and B = 2 twisted pairs of #24 wire, 8 twists per inch. 5 turns each. C = 1 twisted pair of #24 wire, 8 turns. Core - -Stackpole 57-9074-11, Indiana General F624-19Q1 or equivalent.

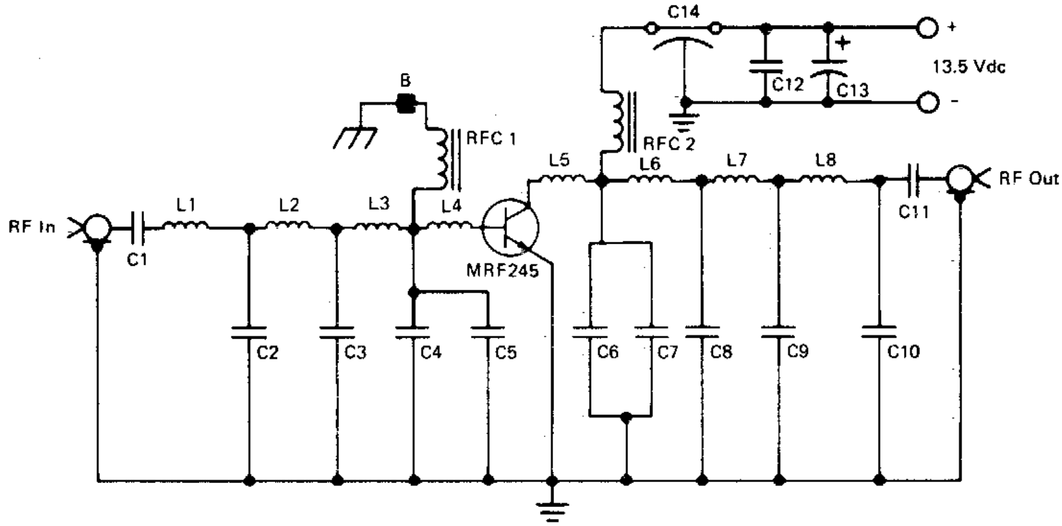
Q1 – 2N6367
Q2, Q3 – 2N6368

D1 – 1N4001
D2 – 1N4997

J1, J2 – BNC connectors

Fig. 73-4

SINGLE-DEVICE, 80 W, 50 Ohm VHF AMPLIFIER



C1, 11 – 500 pF Dipped mica
C2, 9 – 10 pF UNELCO
C3 – 60 pF UNELCO
C4, 5 – 250 pF UNELCO
C6, 7 – 250 pF UNELCO
C8 – 80 pF UNELCO
C10 – 40 pF UNELCO
C12 – 0.1 μF Erie Redcap
C13 – 1 μF Tantalum
C14 – 680 pF Allen Bradley Feed-Thru

RFC 1 – 0.15 μH Molded choke
RFC 2 – 10 T NO. 18 AWG Enameled Wire, 1/4" I.D.
B – Ferroxcube Bead 56-590-65, 3 Beads

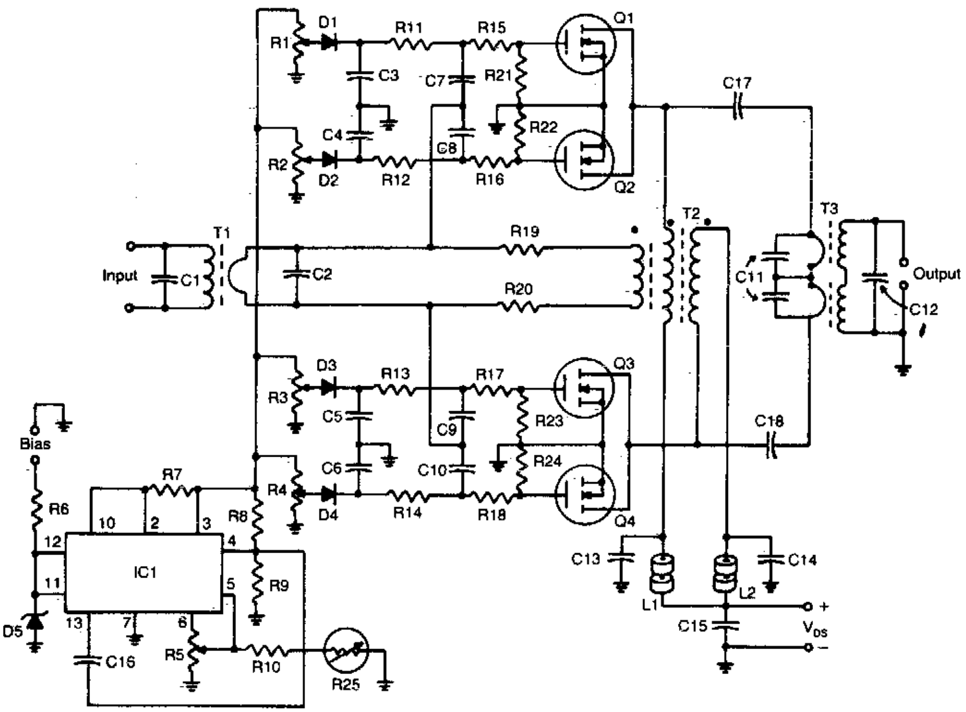
L1 – 1.2 X 0.3 cm Airline Inductor
L2 – 3.5 X 0.3 cm Airline Inductor
L3 – 4.0 X 0.3 cm Airline Inductor
L4, L5 – 0.3 X 0.3 cm Airline Inductor
L6 – 2.7 X 0.3 cm Airline Inductor
L7 – 0.8 X 0.3 cm Airline Inductor
L8 – 3.0 X 0.3 cm Airline Inductor
Board: G10, $\epsilon_r \approx 5$, t = 0.16 cm, 57 gm, Copper-Clad
connectors = BNC

Fig. 73-5

Circuit Notes

The amplifier uses a single MRF245 and provides 80 W with 9.4 dB gain across the 143 to 156 MHz band.

600 W RF POWER AMPLIFIER



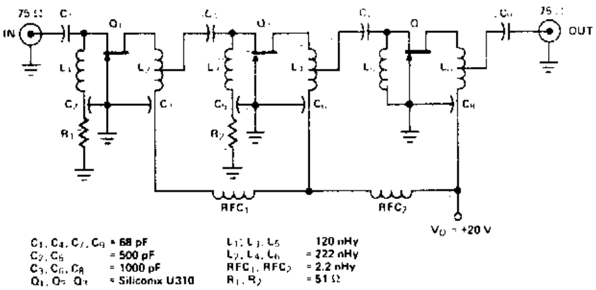
R1-R5—10 kΩ trimpot
R6—1.0 kΩ/1.0W
R7—10 Ω
R8—2.0 kΩ
R9,R21-R24—10 kΩ
R10—8.2 kΩ
R11-R14—100 Ω
R15-R18—1.0 Ω
R19-R20—10 Ω/2.0 W Carbon
R25—thermistor, 10 kΩ (25°C), 2.5 kΩ (75°C)
C1—not used
C2—820 pF ceramic chip
C3-C6, C13,C14—0.1 μF ceramic
C7-C10—0.1 μF ceramic chip
C11—1200 pF each, 680 pF mica in parallel with an Arco 469 variable or three or more smaller value mica capacitors in parallel
C12—not used
C15—10 μF, 100 V electrolytic
C16—1000 pF ceramic
C17,C18—two 0.1 μF, 100 V ceramic each, (ATC 200/823 or equivalent)
D1-D4—IN4148
D5—28 V zener, IN5362 or equivalent
L1,L2—Two Fair-Rite 2673021801 ferrite beads each or equivalent 4.0 μH
T1-T3—see text
Q1-Q4—MRF150
IC1—MC1723CP
All resistors are 0.5W carbon or metal film unless otherwise designated.

Fig. 73-6

Circuit Notes

A unique push-pull parallel circuit. It uses four MRF150 RF power FETs paralleled at relatively high power levels. Supply voltages of 40 to 50 Vdc can be used, depending on linearity requirements. The bias for each device is independently adjustable; therefore, no matching is required for the gate threshold voltages.

WIDEBAND UHF AMPLIFIER WITH HIGH-PERFORMANCE FETs

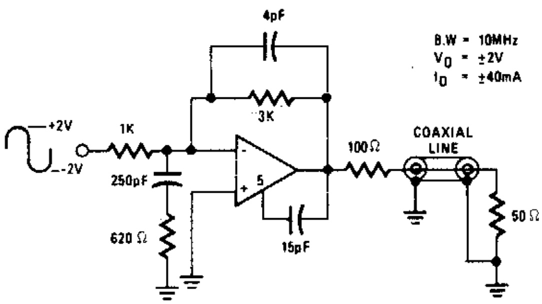


Circuit Notes

The amplifier circuit is designed for 225 MHz center frequency, 1 dB bandwidth of 50 MHz, low input VSWR in a 75-ohm system, and 24 dB gain. Three stages of U310 FETs are used in a straight forward design.

Fig. 73-7

10 MHz COAXIAL LINE DRIVER

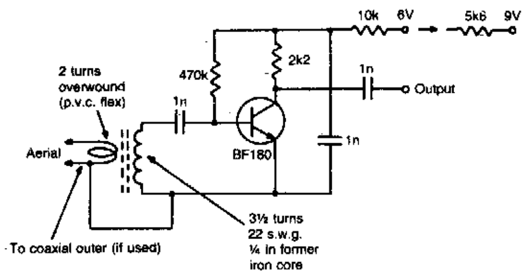


Circuit Notes

The circuit will find excellent usage in high frequency line driving systems that require wide-power bandwidths at high output current levels. (IC=HA2530) The bandwidth of the circuit is limited only by the single pole response of the feedback components; namely f(−3 dB) = ½ π RfCf. As such, the response is flat with no peaking and yields minimum distortion.

Fig. 73-8

VHF PREAMPLIFIER

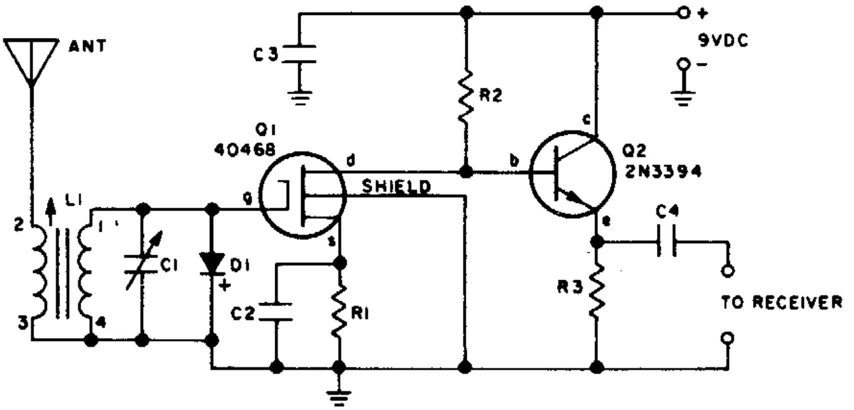


Circuit Notes

This simple circuit gives 15 dB gain and can be mounted on 1 in²PCB. Coil data is given for 85 to 95 MHz. For other frequencies modify coil as required.

Fig. 73-9

SHORTWAVE FET BOOSTER



PARTS LIST FOR SWL'S FET BOOSTER

C1—365-pF tuning capacitor
C2, C3—0.05-uF, 25-VDC capacitor
C4—470-pF, 25-VDC capacitor
D1—1N914 diode
L1—Antenna coil: 1.7-5.5 KHz use Miller B-5495A, 5.5-15 MHz use Miller C-5495A, 12-36 MHz use Miller D-5495-A
Q1—RCA 40468 FET transistor (Do not substitute)
Q2—2N3394 npn transistor
R1—470-ohm, ½-watt resistor
R2—2400-ohm, ½-watt resistor
R3—4700-ohm, ½-watt resistor

Fig. 73-10

Circuit Notes

This two transistor preselector provides up to 40 dB gain from 3.5 to 30 MHz. Q1 (MOSFET) is sensitive to static charges and must be handled with care.

LOW-NOISE 30 MHz PREAMPLIFIER

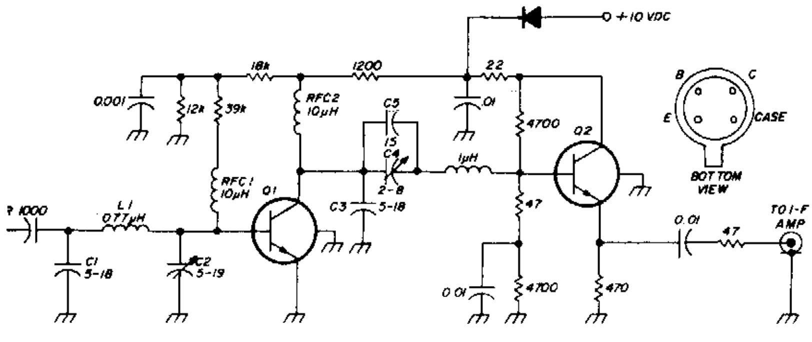


Fig. 73-11

Circuit Notes

Low-noise preamplifier has a noise figure of 1.1 dB at 30 MHz and 3 dB bandwidth of 10 MHz. Gain is 19 dB. Total current drain with a +10 volt supply is 13 mA. All resistors are ¼ watt carbon; bypass capacitors are 50-volt ceramics.

LOW-NOISE BROADBAND AMPLIFIER

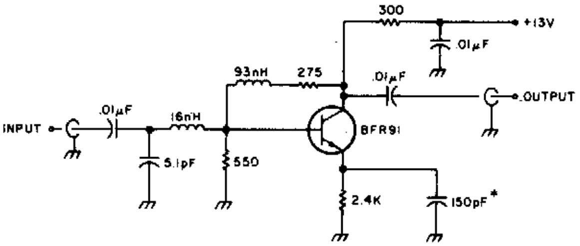


Fig. 73-12

Circuit Notes

The amplifier provides 10 dB of gain from 10-600 MHz and has a 1.5-to-1 match at 50 ohms. The BFR91 has a 1.5 dB noise figures at 500 MHz. The circuit requires 13.5 Vdc at about 13 mA. Keep the leads on the 150 pF emitter bypass capacitor as short as possible. The 16 nH coil is 2.5 turns of #26 enamel wire on the shank of a #40 drill. The 93 nH inductor is 10 turns of the same material.

TWO-METER 10 WATT POWER AMPLIFIER

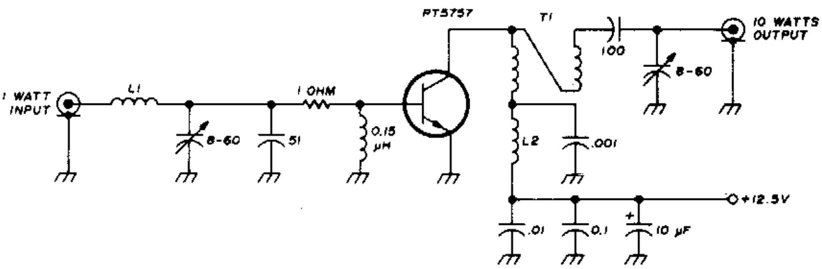
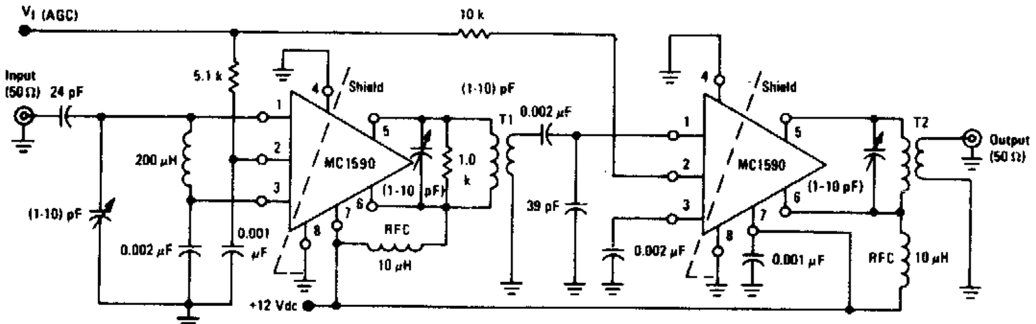


Fig. 73-13

Circuit Notes

This 10-watt, 144-MHz power amplifier uses a TRW PT5757 transistor. L1 is 4 turns of no. 20 enameled, 3/32" ID; L2 is 10 turns of no. 20 enameled, 3/32" ID. Transformer T1 is a 4:1 transmission-line transformer made from a 3" length of twisted pair of no. 20 enameled wire.

TWO-STAGE 60 MHz IF AMPLIFIER
(POWER GAIN ≈ 80 dB, BW ≈ 1.5 MHz)



T1: Primary Winding = 15 Turns, #22 AWG Wire, 1/4" ID Air Core
Secondary Winding = 4 Turns, #22 AWG Wire,
Coefficient of Coupling ≈ 1.0

T2: Primary Winding = 10 Turns, #22 AWG Wire, 1/4" ID Air Core
Secondary Winding = 2 Turns, #22 AWG Wire,
Coefficient of Coupling ≈ 1.0

Fig. 73-14

28 V WIDEBAND AMPLIFIER (3 to 100 MHz)

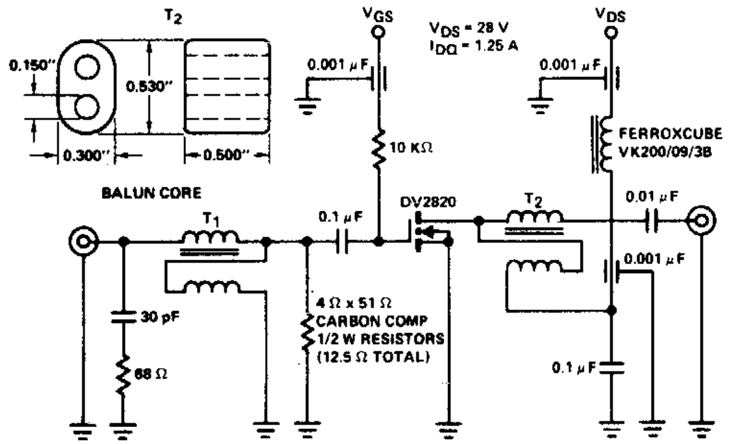
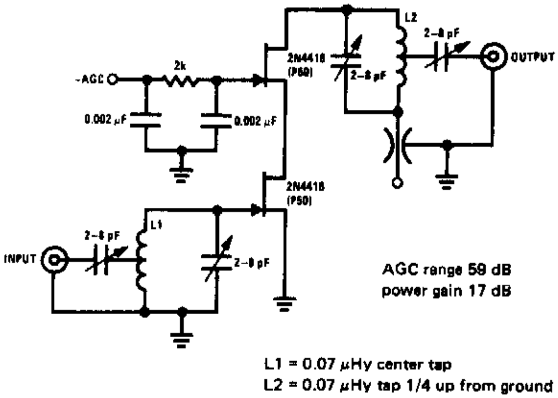


Fig. 73-15

Parts List

T_1, 20 turns 30 Ω, #30 bifilar on micrometals T-50-6 Toroid
T_2, 1 turn of 2-50 Ω coax cables in parallel through 2 balun cores stackpole #57-9130 µo = 125

200 MHz CASCODE AMPLIFIER

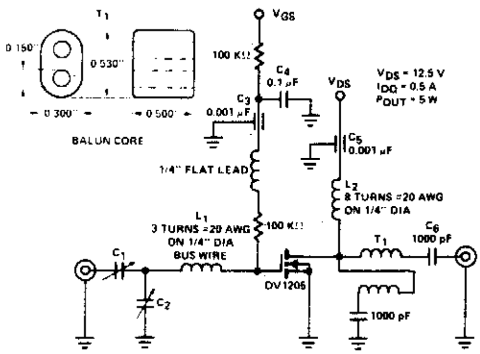


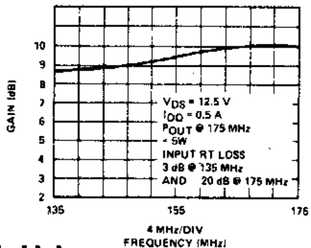
Circuit Notes

This 200 MHz JFET cascode circuit features low cross-modulation, large signal handling ability, no neutralization, and AGC controlled by biasing the upper cascode JFET. The only special requirement of this circuit is that I_{DSS} of the upper unit must be greater than that of the lower unit.

Fig. 73-16

135-175 MHz AMPLIFIER



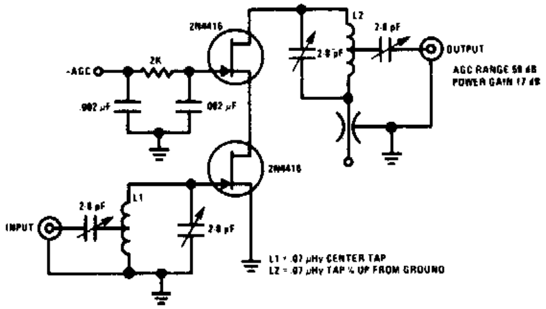


Parts List

C_1, C_2 ARCO #462, 2 to 80 pF, trimmer capacitors
L_1, 3 turns buss wire #20 AWG on 1/4" diameter
L_2, 8 turns #20 AWG on 1/4" diameter
T_1, 1 turn of 25 Ω coax on 2 balun cores.
Stackpole #57-0973 $\mu o = 35$.

Fig. 73-17

200 MHz CASCODE AMPLIFIER

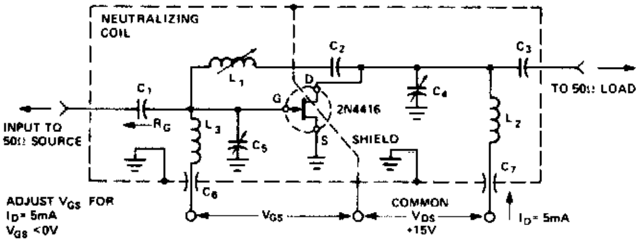


Circuit Notes

This 200 MHz JFET cascode circuit features low cross-modulation, large signal handling ability, no neutralization, and AGC controlled by biasing the upper cascode JFET. The only special requirement of this circuit is that I_{DSS} of the upper unit must be greater than that of the lower unit.

Fig. 73-18

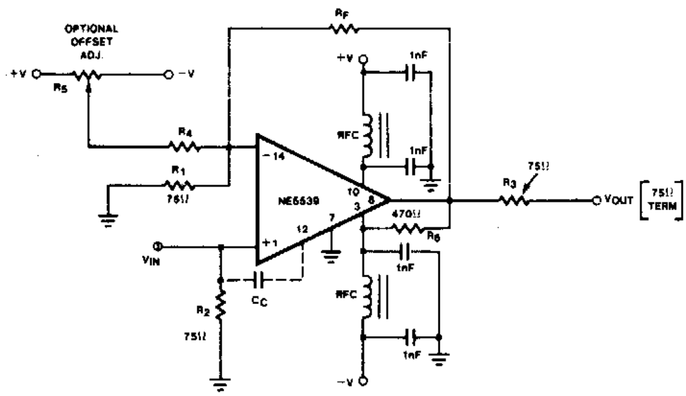
100 MHz AND 400 MHz NEUTRALIZED COMMON SOURCE AMPLIFIER



REFERENCE DESIGNATION	VALUE 100MHz	VALUE 400MHz
C_1	7.0pF	1.8pF
C_2	1000pF	27pF
C_3	3.0pF	1.0pF
C_4	1.0–12pF	0.8–8pF
C_5	1.0–12pF	0.8–8pF
C_6	0.0015μF	0.001μF
C_7	0.0015μF	0.001μF
L_1	3.0μH	0.2μH
L_2	0.25μH	0.03μH
L_3	0.14μH	0.022μH
Typ NF	1.2dB	2.4dB
Typ G_{ps}	21dB	12dB

Fig. 73-19

ULTRA HIGH FREQUENCY AMPLIFIER



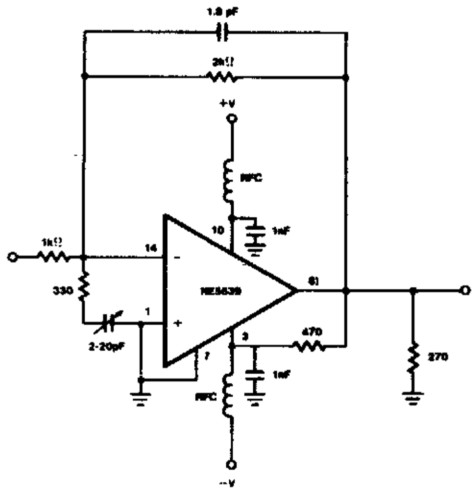
R_1 = 75Ω 5% CARBON
R_2 = 75Ω 5% CARBON
R_3 = 75Ω 5% CARBON
R_4 = 36K 5% CARBON

R_5 = 20K TRIMPOT (CERMET)
R_F = 1.5K (28dB GAIN)
R_6 = 470Ω 5% CARBON

RFC 3T # 28 BUSSWIRE ON
FERROXCUBE VK 200 09/3B CORE
BYPASS CAPACITORS
1nF CERAMIC
(MEPCO OR EQUIV.)

Fig. 73-20

UHF AMPLIFIER WITH INVERTING GAIN OF 2 AND LAG-LEAD COMPENSATION (GAIN BANDWIDTH PRODUCT 350 MHz)



NOTE
Resistors — 1/4 watt carbon.
RFC-3T #26 bus wire on Ferroxcube VK200 09/3B wideband threaded core.

Fig. 73-21

TRANSISTORIZED Q-MULTIPLIER FOR USE WITH IFS IN THE 1400 kHz RANGE

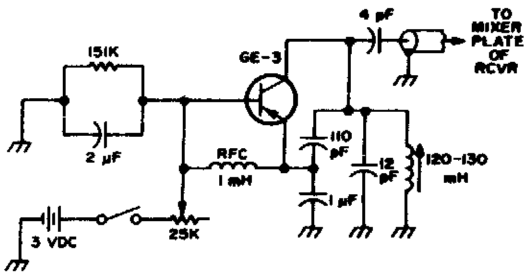
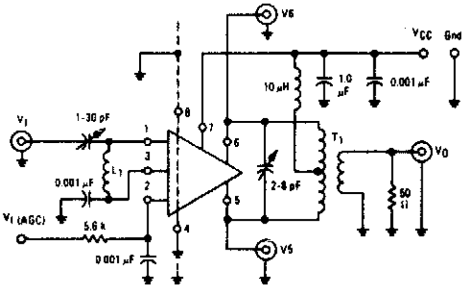


Fig. 73-22

60 MHz AMPLIFIER

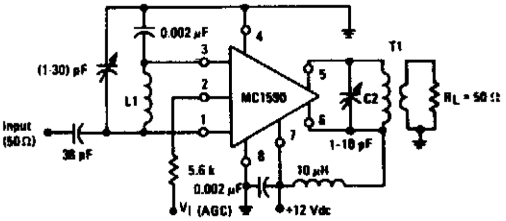


L1 7 Turns, # 22 AWG Wire on 5/16" Dia. Form, 5/8" Long

T1: Close Wound Over 1/4" Form
Primary Winding = 16 Turns # 26 AWG, Center Tapped
Secondary Winding = 2 Turns # 26 AWG.

Fig. 73-23

30 MHz AMPLIFIER (POWER GAIN = 50 dB, BW ≈ 1.0 MHz)

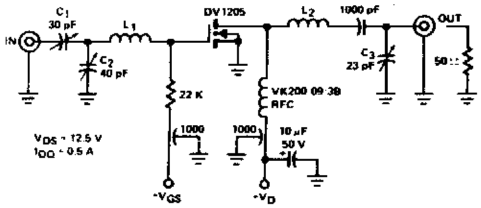


L1 = 12 Turns #22 AWG Wire on a Toroid Core, (T37-6 Micro Metal or Equiv)

T1: Primary = 17 Turns #20 AWG Wire on a Toroid Core, (T44-6 Micro Metal or Equiv)
Secondary = 2 Turns #20 AWG Wire

Fig. 73-24

TWO METER AMPLIFIER, 5 W OUTPUT



Parts List

L_1, 60 nHy 4T #22 AWG close wound 0.125" I.D.
L_2, 54 nHy 3 1/2T #22 AWG close wound 0.125" I.D.
C_1, C_2, C_3, ARCO #462 5–80 pF

Fig. 73-25

80 MHz CASCODE AMPLIFIER

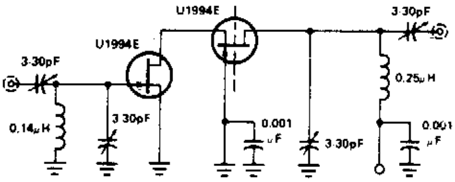
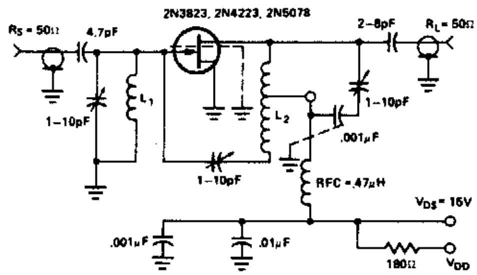


Fig. 73-26

200 MHz NEUTRALIZED COMMON SOURCE AMPLIFIER

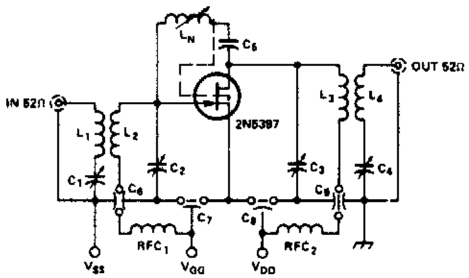


L_1 1-1/2 turns, #20 tinned wire, 1/4 ID, Length = 3/8"
L_2 3-1/2 turns, #18 tinned wire, 3/8" ID, Length = 1/2"
Tapped at 1- 1/4 turns from drain

NF Typ 1.5dB
G_{ps} Typ 18dB
V_{DS} = +15V
V_{GS} = 0

Fig. 73-27

450 MHz COMMON-SOURCE AMPLIFIER



C_{1-4} – 0.8 – 12pF Johanson type 2950
C_5 – 40pF DMS silver mica
C_{6-9} – 1000pF Allen-Bradley type FA5C
L_1 – 1.4" long; #22 enamel spaced 0.1" from L_2
L_2 – 1.1" long; #16 solid copper
L_3 – 1.3" long; #16 solid copper
L_4 – 1.4" long; #22 enamel spaced 0.3" from L_3
$RFC_{1,2}$ – 0.15μH Delevan type 1537-00
L_N – 3T, #22 enamel; 0.25" diam. ceramic form; aluminum slug, low loss

NF Typ 2.8dB
G_{ps} Typ 18dB
V_{DG} = +10V
I_D = 10mA

Fig. 73-28

74

RF Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

500 MHz OSCILLATOR

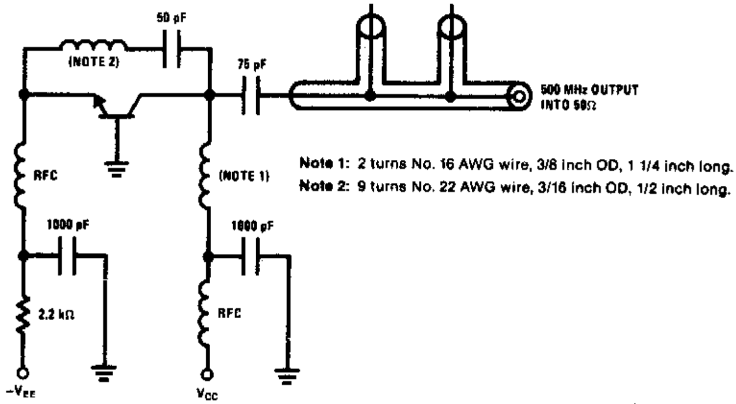
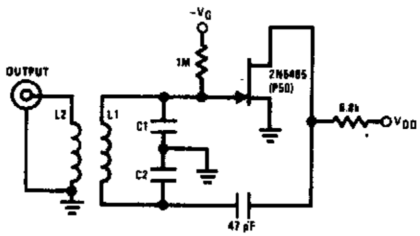


Fig. 74-1

LOW DISTORTION OSCILLATOR



20 MHz oscillator values

C1 ≅ 700 pF	L1 = 1.3 μH
C2 = 75 pF	L2 = 10T 3/8" dia 3/4" long
V_{DD} = 16V	I_D = 1 mA

20 MHz oscillator performance

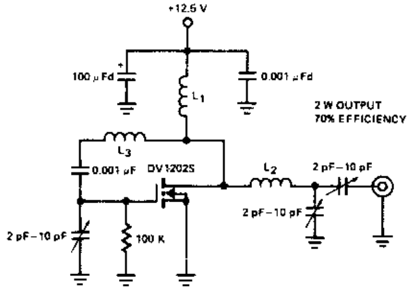
Low distortion 20 MHz osc
2nd harmonic − 60 dB
3rd harmonic > −70 dB

Fig. 74-2

Circuit Notes

The 2N5485 JFET is capable of oscillating in a circuit where harmonic distortion is very low. The JFET local oscillator is excellent when a low harmonic content is required for a good mixer circuit.

400 MHz OSCILLATOR



Parts List

L_1—8 turns #22 closewound on 1/4" diameter
L_2—1/2 inch #16 wire
L_3—1 inch #16 wire

Fig. 74-3

1.0 MHz OSCILLATOR

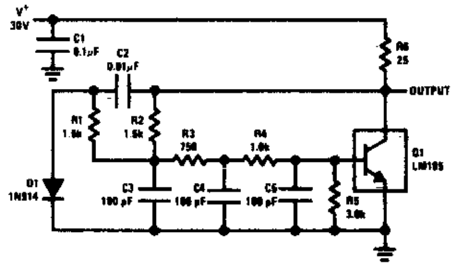


Fig. 74-5

2 MHz OSCILLATOR

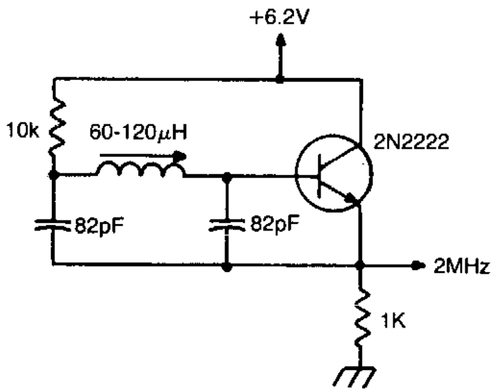


Fig. 74-4

Circuit Notes

Miller 9055 miniature slugtuned coil; all resistors 1/4W 5%; all caps min. 25 V ceramic.

HARTLEY OSCILLATOR

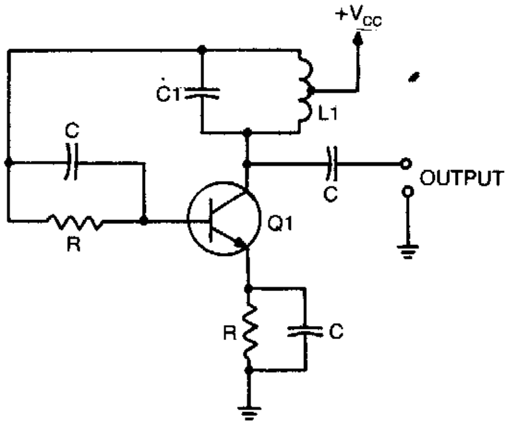
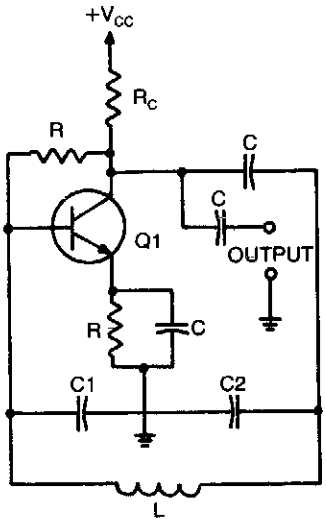


Fig. 74-6

Circuit Notes

Resonant frequency is ½ π $\sqrt{L1C1}$.

COLPITTS OSCILLATOR

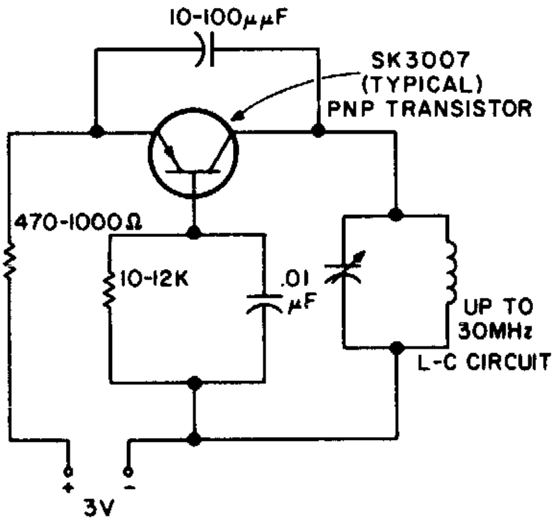


Circuit Notes

When calculating its resonant frequency, use C1C2/C1+C2 for the total capacitance of the L-C circuit.

Fig. 74-7

RF OSCILLATOR



Circuit Notes

This rf oscillator is useful up to 30 MHz. An SK 3007 PNP transistor is recommended.

Fig. 74-8

75

Remote Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Radio Control Receiver/Decoder
Carrier Operated Relay
Remote Control Servo System
Tone-Actuated Relay
Radio Control Motor Speed Controller
Remote On-Off Switch
Automatic Turn Off for TV Set

RADIO CONTROL RECEIVER/DECODER

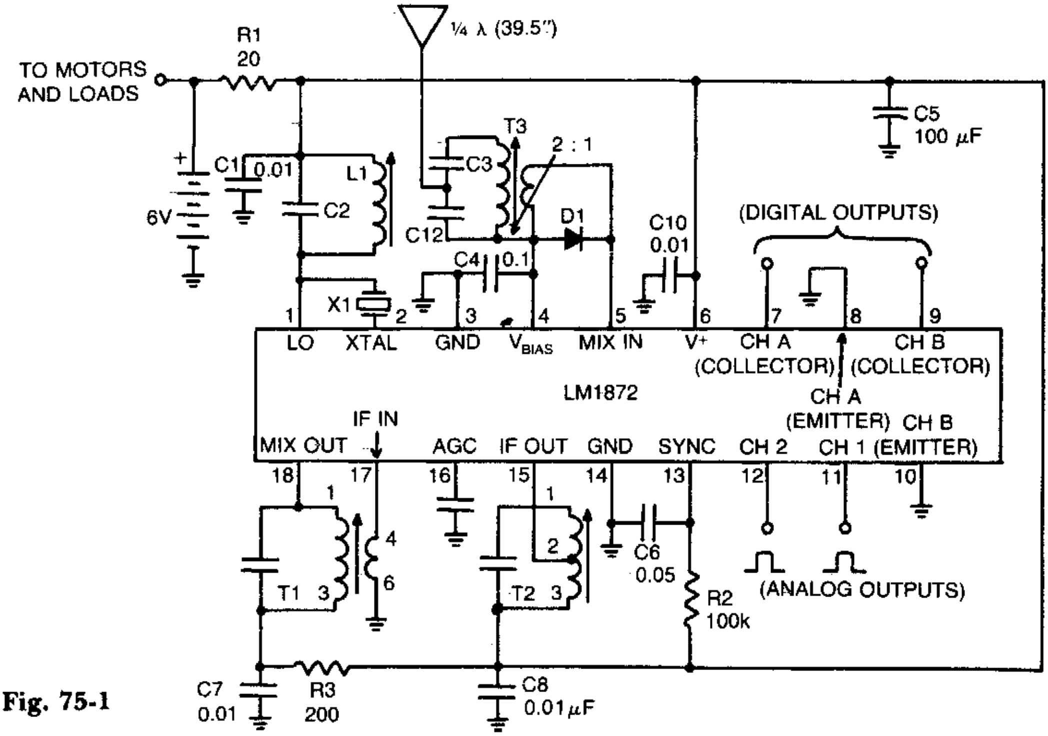
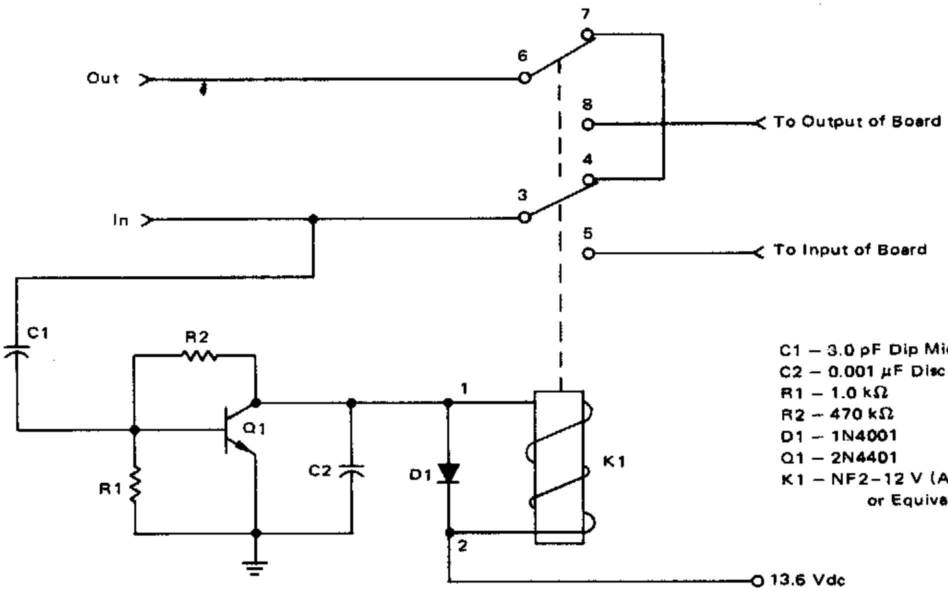


Fig. 75-1

R1 - Motor decoupling
R2 - Sync timer; $R2 = \frac{t\ SYNC}{0.7\ C6}$, R2 ≤ 470k
R3 - Mixer decoupling
C1 - LO bypass; optional
C2 - LO tank; C2 = 22 pF @ 72 MHz
C3 - Ant. input tank; C3 = 24 pF @ 72 MHz
C4 - V_{BIAS} bypass
C5 - Motor decoupling
C6 - Sync timer; $C6 = \frac{t\ SYNC}{0.7\ R2}$, C6 + 0.5 µF
C7 - Mixer decouple; 0.01 µF ≤ C7 ≤ 1 µF
C8 - AGC
C9 - IF bypass; optional
C10 - V+ bypass; 0.01 µF ≤ C10 ≤ 0.1 µF
C12 - Ant. input tank; C12 = 160 pF @ 72 MHz
L1 - LO coil
Toko* 10k type (KENC) 4T; 0.2 µH @ 72 MHz
L1 could be made a fixed coil, if desired.
T1 - 455 kHz mixer transformer
Toko* 10 EZC type (RMC-502182), Qu = 110
Pin 1-2, 82T; pin 2-3, 82T
Pin 1-3, 164T; pin 4-6, 30T
T2 - 455 kHz IF transformer
Toko* 10 EZC type (RMC-502503), Qu = 110
Pin 1-2, 82T; pin 2-3, 8T
T3 - Ant. input transformer
Toko 10k type (KENC), 4T sec. & 2T pri. of 0.2 µH@ 72 MHz
X1 - 5th overtone crystal, parallel-mode, 72 MHz
D1 - Electrostatic discharge (ESD) protection
* Toko America, Inc.
5520 West Touhy Ave.
Skokie, Ill. 60077
(312)677-3640 Tlx: 72-4372

CARRIER OPERATED RELAY



C1 – 3.0 pF Dip Mica
C2 – 0.001 µF Disc
R1 – 1.0 kΩ
R2 – 470 kΩ
D1 – 1N4001
Q1 – 2N4401
K1 – NF2–12 V (Arrow-M) or Equivalent

Fig. 75-2

REMOTE CONTROL SERVO SYSTEM

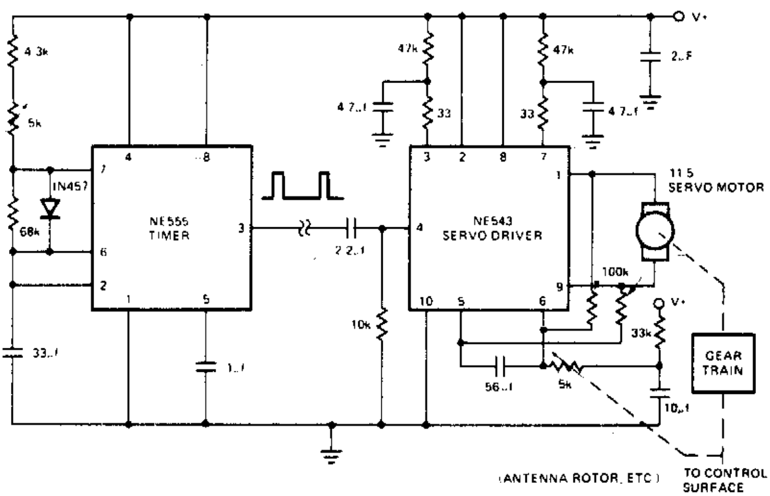
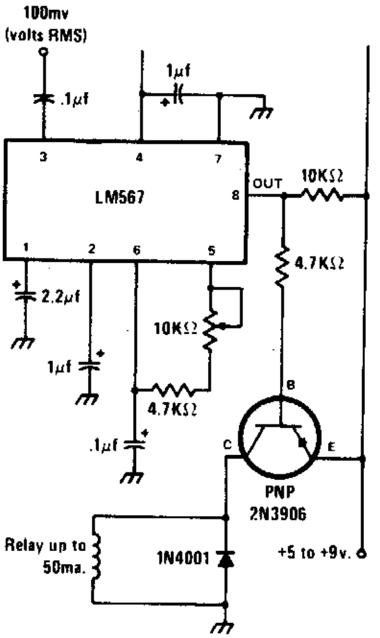


Fig. 75-3

TONE-ACTUATED RELAY



Circuit Notes

The circuit is built around the LM567 tone decoder IC that requires about 100 millivolts at its operating frequency. The frequency is set by a 10 K variable resistor and can be between 700 and 1500 Hz. When a tone at the set frequency is present, the 567's output goes low to energize a relay through a 2N3906 PNP transistor.

Fig. 75-4

RADIO CONTROL MOTOR SPEED CONTROLLER

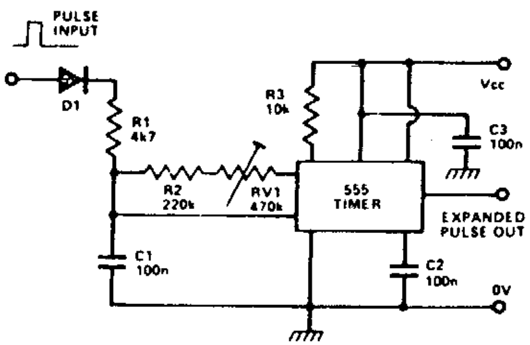


Fig. 75-5

REMOTE ON-OFF SWITCH

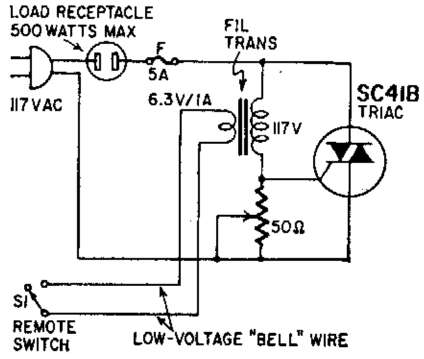


Fig. 75-6

Circuit Notes

This circuit provides power control without running line-voltage switch leads. The primary of a 6-volt filament transformer is connected between the gate and one of the main terminals of a triac. The secondary is connected to the remote switch through ordinary low-voltage line. With switch open, transformer blocks gate current, prevents the triac from firing and applying power to the equipment. Closing the switch short-circuits the secondary, causing the transformer to saturate and trigger the triac.

AUTOMATIC TURN OFF FOR TV SET

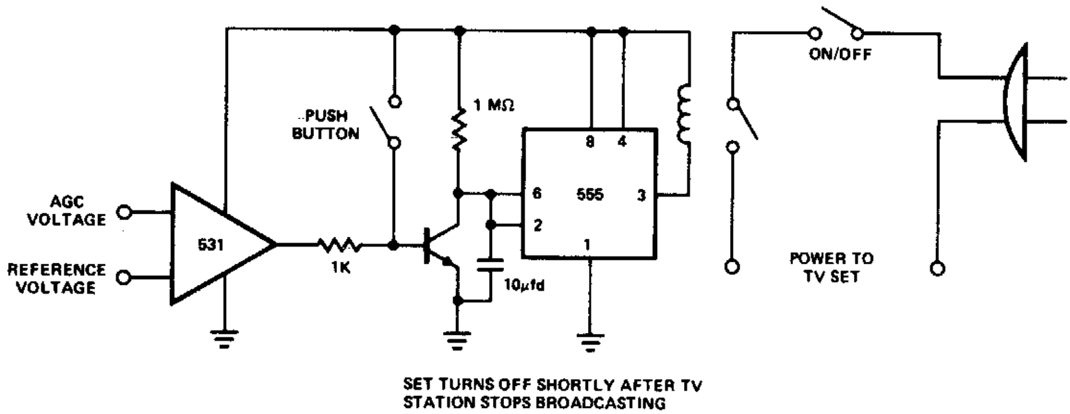


Fig. 75-7

76

Safety and Security Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Tarry Light
Ground Tester
Ground-Fault Interrupter
Single Source Emergency Lighting System
Power Failure Alarm
Ac Hot Wire Probe
Power Failure Detector
Power-Failure Alarm
Electronic Combination Lock

TARRY LIGHT

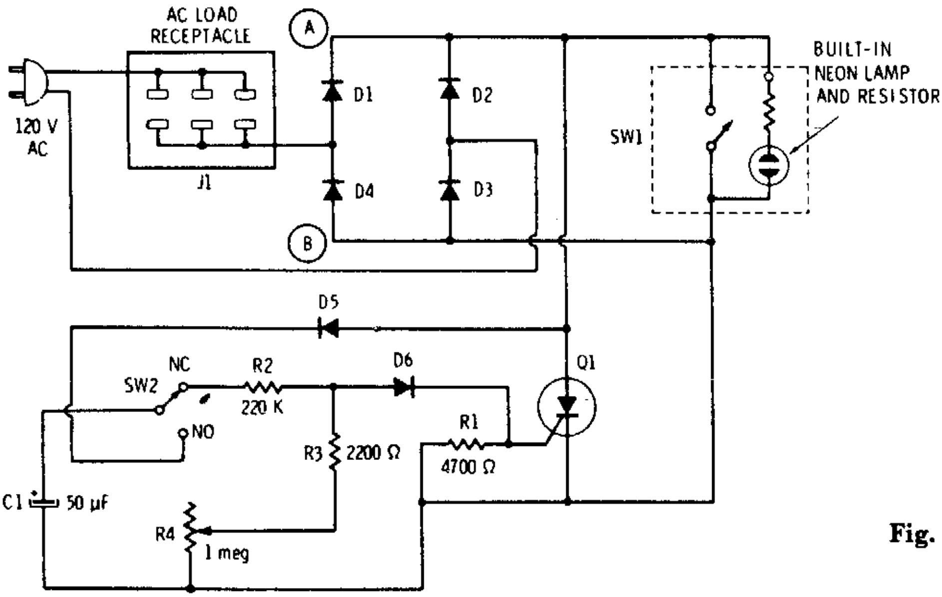
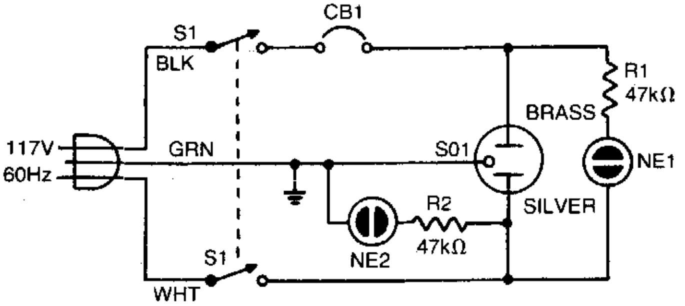


Fig. 76-1

Circuit Notes

The push button and potentiometer initiate a time delay that turns a light on then automatically turns it off again after a predetermined time. The potentiometer can be set for a delay of a few seconds to just under three minutes. When the push-button switch SW2 is pressed, capacitor C1 gets charged through D5 to the full dc voltage developed by the diode bridge. When the button is released, the charged capacitor is connected across the series combination of R2, R3, and potentiometer R4 whose setting determines the total resistance and thereby sets the time it takes for the capacitor to discharge. A steering diode, D6, connected to the junction of R2 and R3, and potentiometer R4 whose setting determines The total resistance and thereby sets the time it takes for the capacitor to discharge. Diode, D6 picks off a portion of this decaying dc voltage and applies it to the gates terminal of Q1, the SCR, triggering it into a conductive state. This SCR will remain on as long as there is sufficient voltage on its gate. As soon as this voltage decays below the minimum holding voltage of the SCR, it will turn off on the next line alternation.

GROUND TESTER



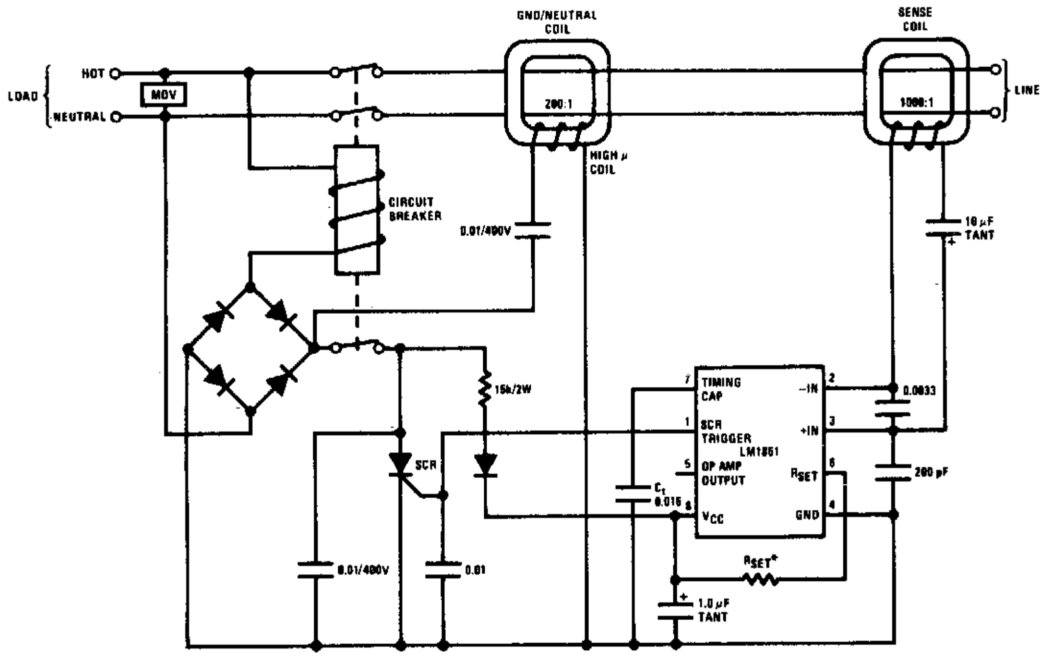
Parts list
S2—DPDT Switch
CB1—10A fuse or circuit breaker
S01—Radio Shack 61-2760, 3 terminal socket
R1, R2, R3—47kΩ, ½W
NE1, NE2, NE3—GE NE-2
S1—SPDT, Lafayette 34P0238V

Fig. 76-2

Circuit Notes

This circuit checks the reliability of appliances so that the equipment may be used safely. The test circuit must be plugged into a properly wired three terminal wall outlet. When a two-lead or three-lead appliance is plugged into circuit outlet S01, neon lamps NE1 and NE2 will light if the appliance is safe. If neon NE2 is lit the appliance is dangerous, because the neutral lead is 110 Vac above ground.

GROUND-FAULT INTERRUPTER (120 Hz NEUTRAL TRANSFORMER APPROACH)



* Adjust R_{SET} for desired sensitivity

Fig. 76-3

SINGLE SOURCE EMERGENCY LIGHTING SYSTEM

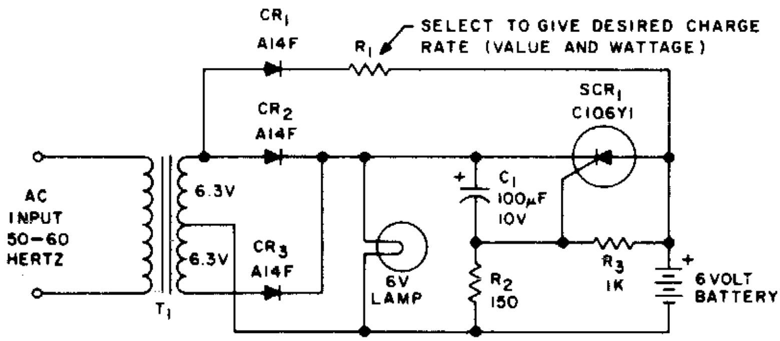
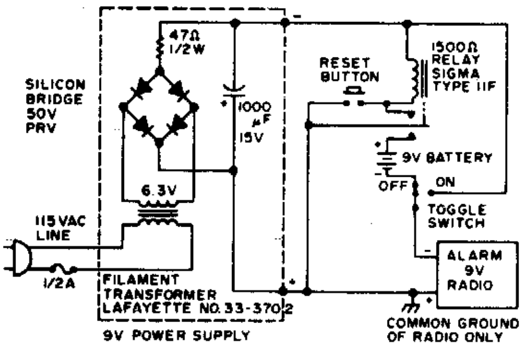


Fig. 76-4

Circuit Notes

This emergency lighting system maintains a 6 volt battery at full charge and switches automatically from the ac supply to the battery.

POWER FAILURE ALARM

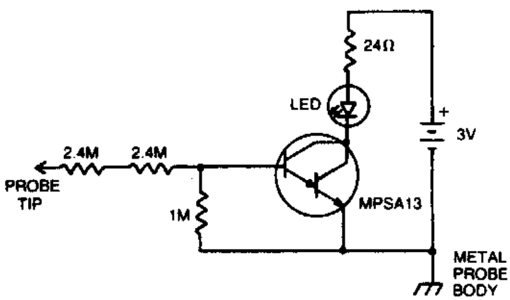


Circuit Notes

If the power fails, the radio alarm goes on. No loud siren, bell, or whistle. Even if the power is restored, the alarm stays on until RESET button is pushed.

Fig. 76-5

AC HOT WIRE PROBE



Circuit Notes

Insert the probe tip into either terminal of an ac outlet and hold the probe body against anything that the circuit ground is connected to. The LED will glow when the hot terminal is touched. Two 2.4 M resistors are used in the probe tip for safety (redundancy) reasons.

Fig. 76-6

POWER FAILURE DETECTOR

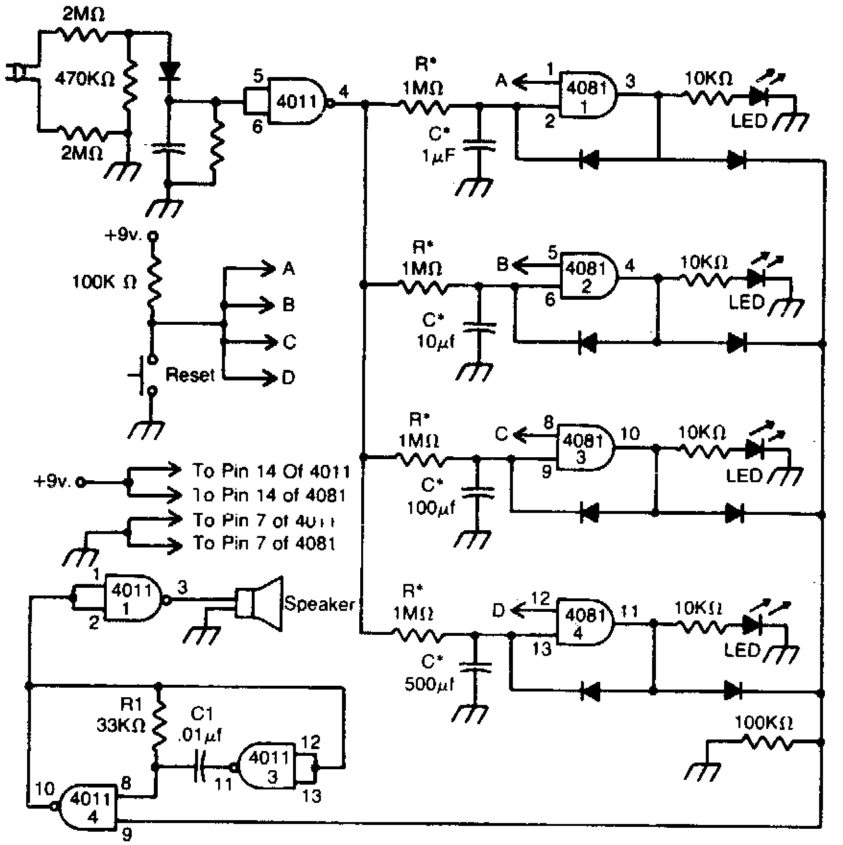


Fig. 76-7

Circuit Notes

This circuit indicates that a power outage occured for 1, 10, 100, and 500 seconds with the values given for R* and C*. After a power failure, the circuit can be reset by pushing the Reset button.

POWER FAILURE ALARM

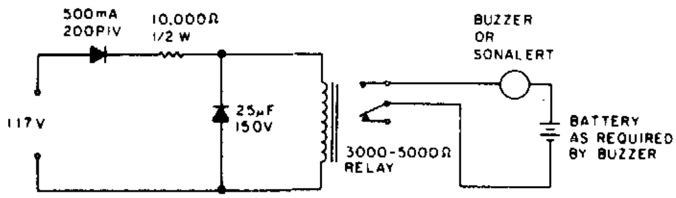


Fig. 76-8

Circuit Notes

While the power is on, the relay is held open, but when the power fails the buzzer-circuit contacts close.

ELECTRONIC COMBINATION LOCK

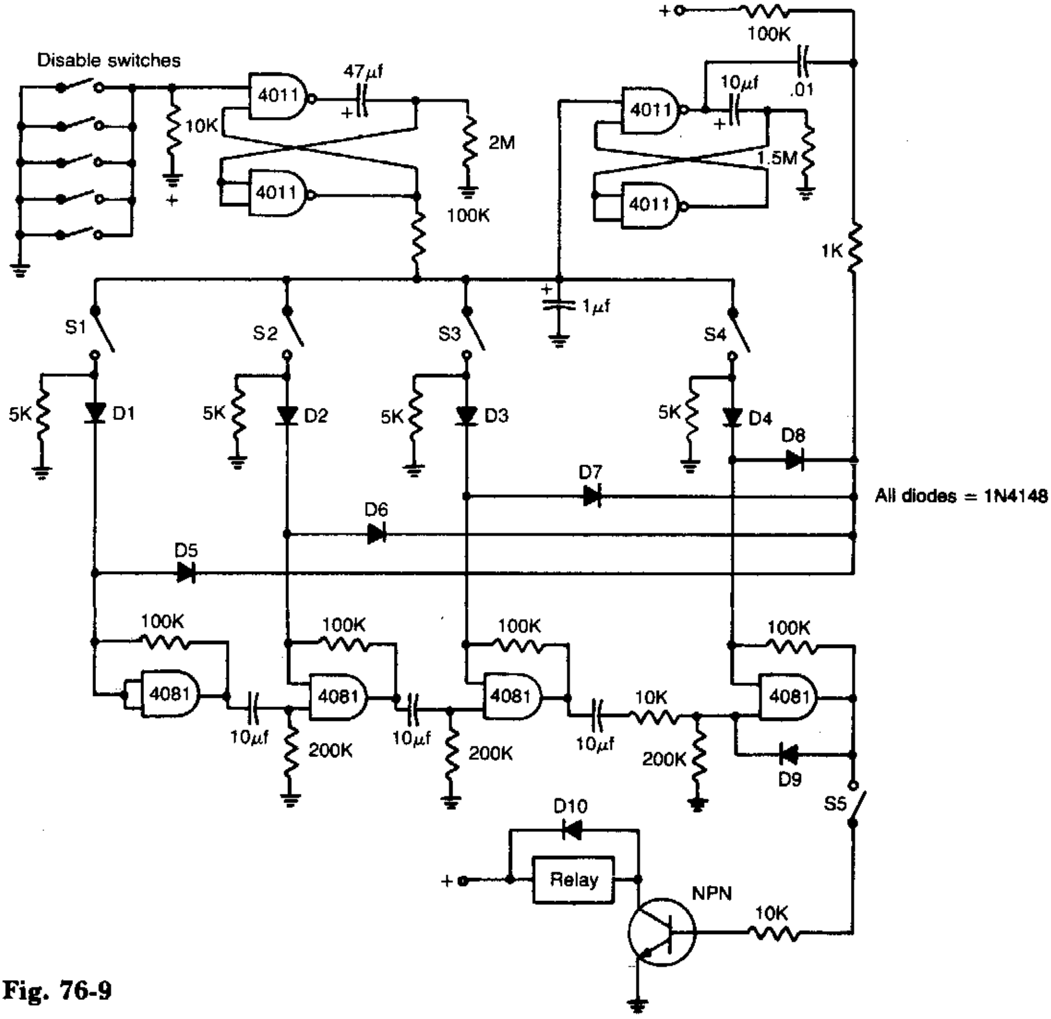


Fig. 76-9

Circuit Notes

Switches S1 through S5 must be operated in rapid sequence to operate the lock. They can be any numbers on a 10-button switch pad. If an incorrect button is pushed, alarm sounds and the circuit is disabled for two minutes.

77

Sample and Hold Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PEAK DETECT AND HOLD

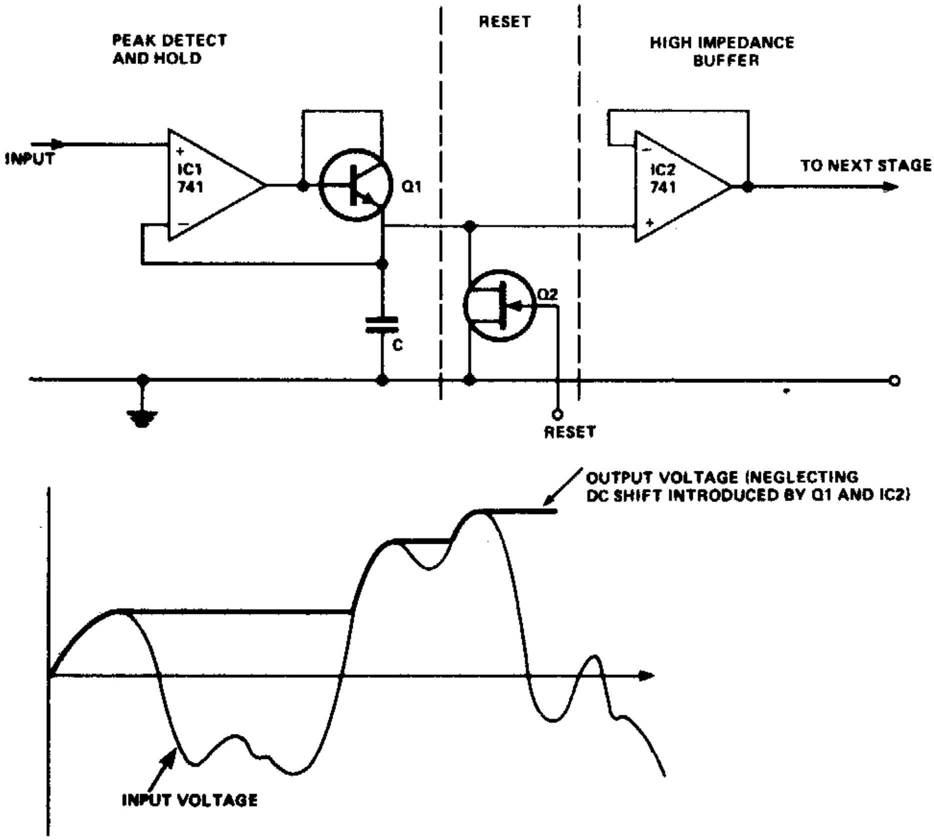
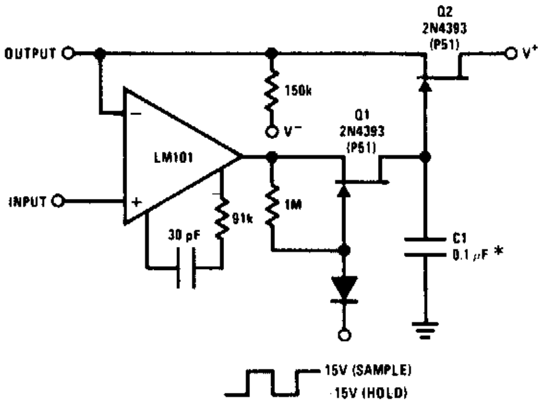


Fig. 77-1

Circuit Notes

If the voltage at the input exceeds the voltage on the capacitor, then the output of the 741 goes positive, the diode conducts, and the capacitor is charged up to the input voltage-forward voltage drop of diode. When the voltage at the input is less than that on the capacitor, the output of the 741 goes negative, and the diode cuts off. To prevent the capacitor from discharging through the input resistance of the next stage, a high input impedance buffer stage (IC2) is used. The circuit can be reset by means of a FET or similar high impedance device connected across the capacitor.

LOW DRIFT SAMPLE AND HOLD



*Polycarbonate dielectric capacitor

Circuit Notes

The JFETs, Q1 and Q2, provide complete buffering to C1, the sample and hold capacitor. During sample, Q1 is turned on and provides a path, $r_{ds(on)}$, for charging C1. During hold, Q1 is turned off, thus leaving Q1 $I_{D(off)}$ (< 100 pA) and Q2 I_{GSS} (< 100 pA) as the only discharge paths. Q2 serves a buffering function so feedback to the LM101 and output current are supplied from its source.

Fig. 77-2

JFET SAMPLE AND HOLD

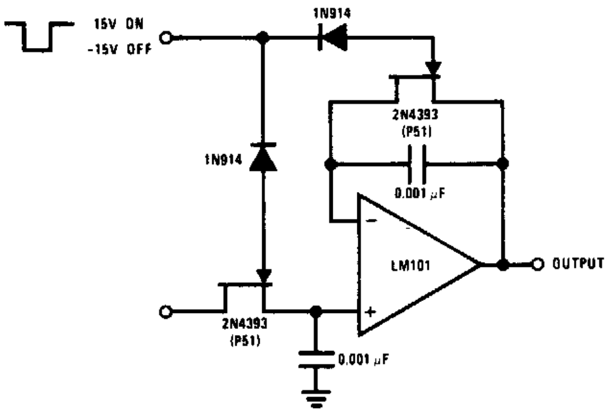


Fig. 77-3

Circuit Notes

The logic voltage is applied simultaneously to the sample and hold JFETs. By matching input impedance and feedback resistance and capacitance, errors due to $r_{ds(on)}$ of the JFETs are minimized.

HIGH SPEED SAMPLE AND HOLD AMPLIFIER

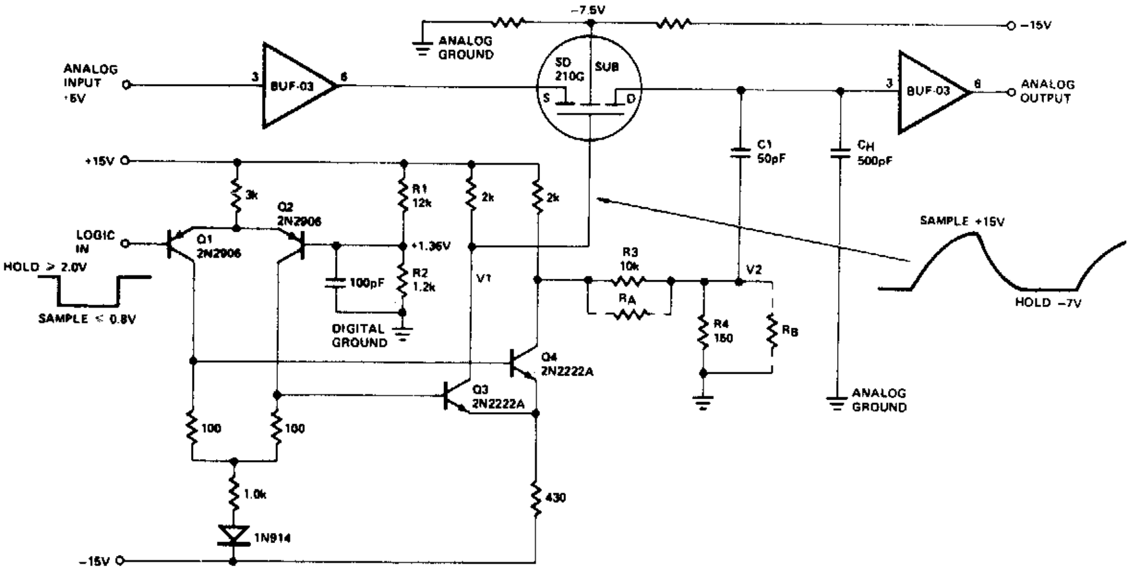


Fig. 77-4

HIGH SPEED SAMPLE AND HOLD

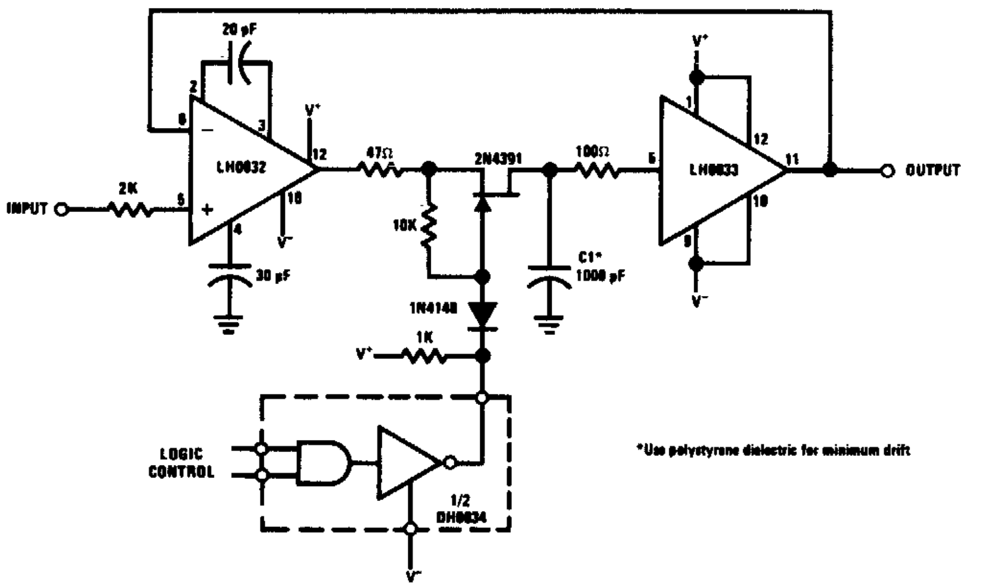


Fig. 77-5

HIGH SPEED SAMPLE AND HOLD

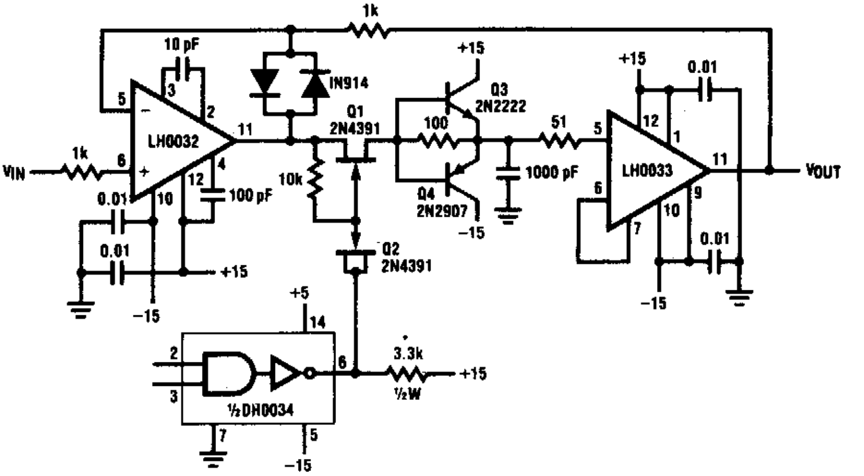


Fig. 77-6

Circuit Notes

This circuit exhibits a 10 V acquisition time of 900 ns to 0.1% accuracy and a droop rate of only 100 μV/ms at 25° C ambient condition. An even faster acquisition time can be obtained using a smaller value hold-capacitor. By decreasing the value from 1000 pF to 220 pF, the acquisition time improves to 500 ns for a 10 V step. However, the droop rate increases to 500 μV/ms.

SAMPLE AND HOLD WITH OFFSET ADJUSTMENT

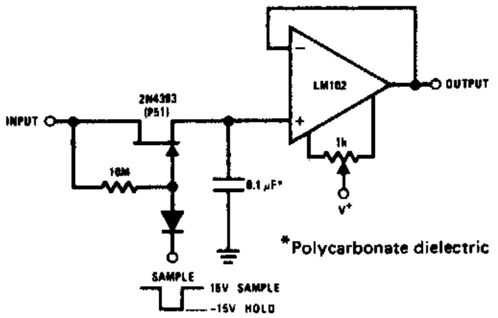


Fig. 77-7

Circuit Notes

The 2N4393 JFET was selected because of its low I_{GSS} (< 100 pA), very low $I_{D(off)}$ (< 100 pA) and low pinchoff voltage. Leakages of this level put the burden of circuit performance on clean, solder-resin free, low leakage circuit layout.

DIFFERENTIAL HOLD

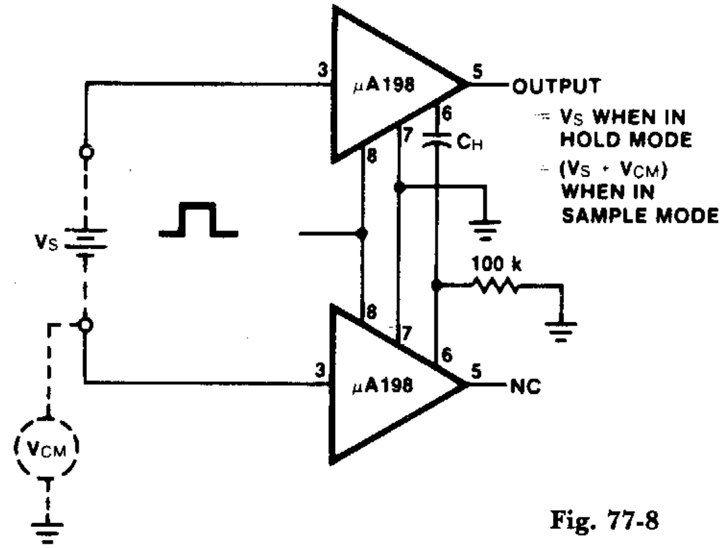
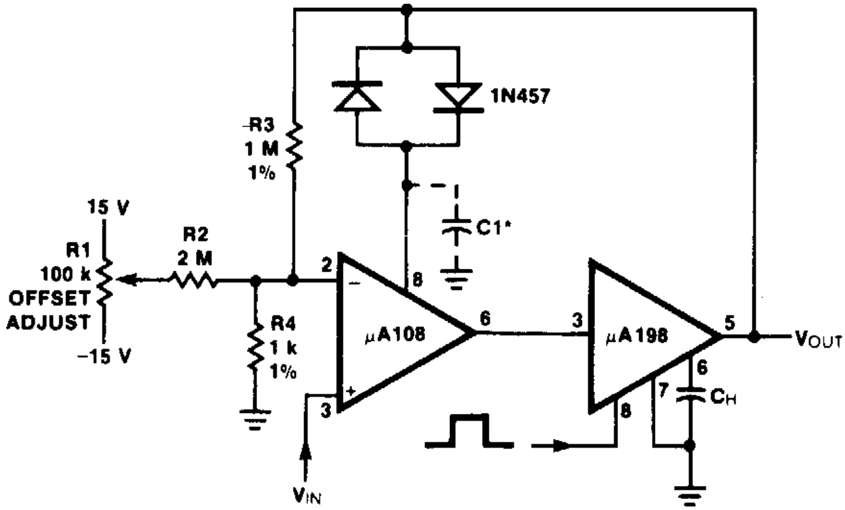


Fig. 77-8

× 1000 SAMPLE AND HOLD



Notes

For lower gains, the μA108 must be frequency compensated

Use $\approx \frac{100}{A_V}$ pF from comp 2 to ground

Fig. 77-9

SAMPLE AND HOLD

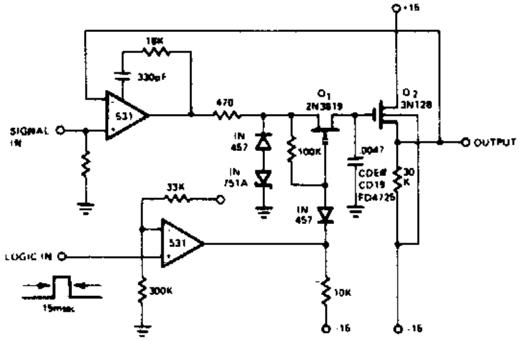
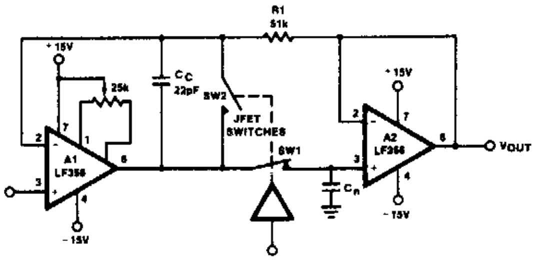


Fig. 77-10

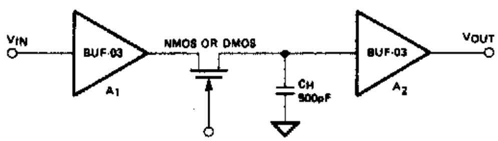
HIGH ACCURACY SAMPLE AND HOLD



- By closing the loop through A2 the V_{OUT} accuracy will be determined uniquely by A1. No V_{OS} adjust required for A2.
- T_A can be estimated by same considerations as previously but, because of the added on propagation delay in the feedback loop (A2) the overshoot is not negligible.
- Overall system slower than fast sample and hold.
- R1, C_C: additional compensation
- Use LF356 for
  - Δ Fast settling time
  - Δ Low V_{OS}

Fig. 77-11

HIGH SPEED SAMPLE AND HOLD



ICHARGE OF BUF-03 IS ±60mA. THEREFORE THE SLEW RATE INTO A 500pF HOLD CAPACITOR WILL BE 120V/μSEC. THUS THE SLEW RATE OF THE SAMPLE AND HOLD CIRCUIT IS LIMITED BY THE CAPACITOR CHARGING TIME.

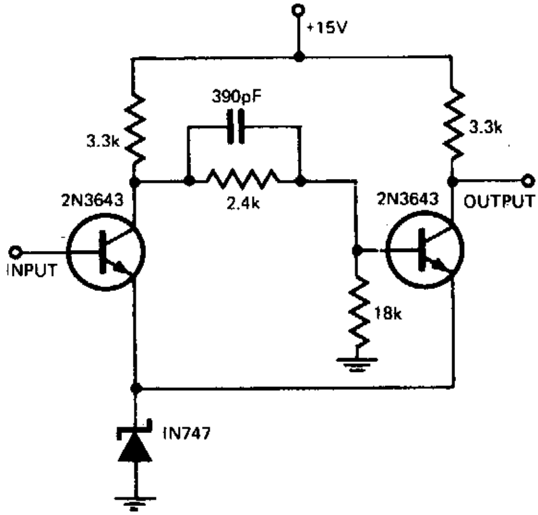
Fig. 77-12

78

Schmitt Triggers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SCHMITT TRIGGER WITHOUT HYSTERESIS



Circuit Notes

By replacing the common-emitter resistor in a conventional Schmitt by a zener diode, the hysteresis normally associated with these circuits is eliminated.

Fig. 78-1

SCHMITT TRIGGER WITH PROGRAMMABLE HYSTERESIS

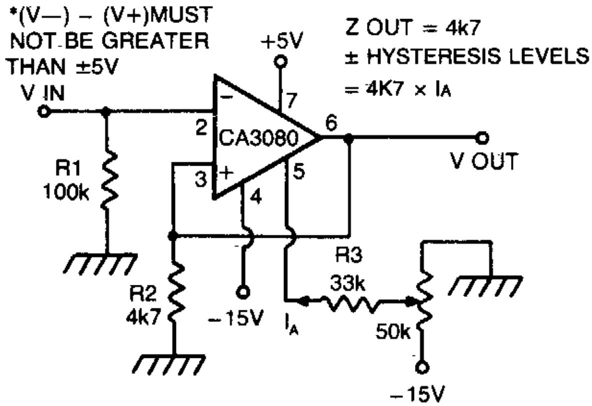


Fig. 78-2

Circuit Notes

CA 3088 is used as a versatile Schmitt trigger. The size of the hysteresis levels is determined by I_A that flows out of the amplifier's output and through R2. Increasing I_A increases hysteresis and vice versa. The positive and negative hysteresis levels are symmetrical about 0 V.

SCHMITT TRIGGER (ZERO CROSSING DETECTOR WITH HYSTERESIS)

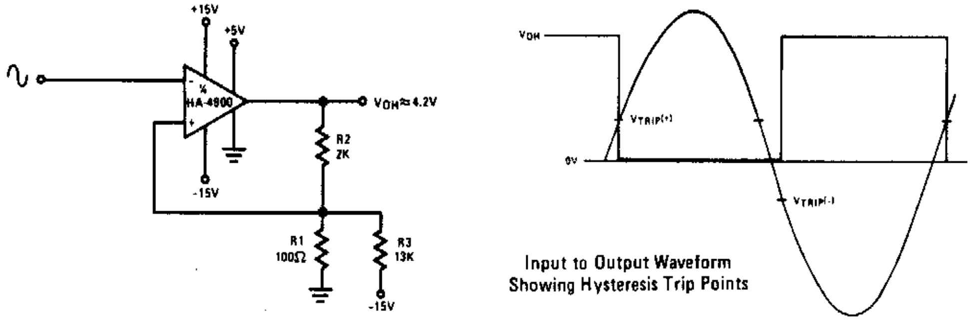


Fig. 78-3

Circuit Notes

This circuit has a 100 mV hysteresis which can be used in applications where very fast transition times are required at the output even though the signal is very slow. The hysteresis loop also reduces false triggering due to noise on the input. The waveforms show the trip points developed by the hysteresis loop.

SCHMITT TRIGGER

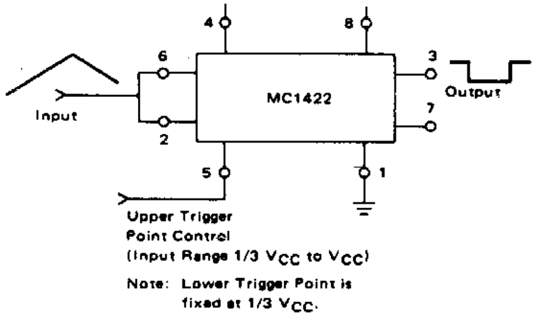


Fig. 78-4

Circuit Notes

The lower trigger point is fixed at ⅓ Vcc, but the upper trigger point is adjustable by means of Pin 5 from ⅓ Vcc to slightly less than Vcc. The Schmitt trigger will operate with input frequencies up to 50 kHz.

79

Smoke and Flame Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PHOTOELECTRIC SMOKE DETECTOR (NON-LATCHING)

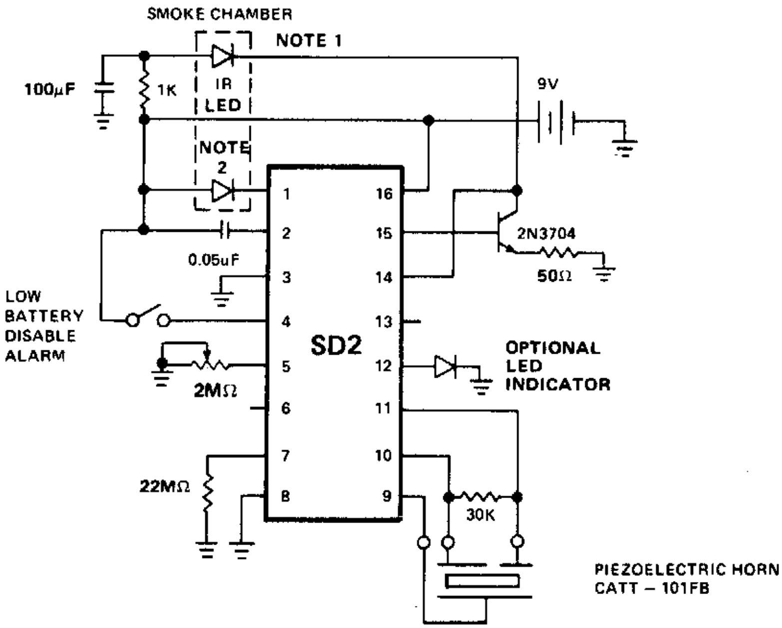


Fig. 79-1

Notes: 1. IR Diode RCA Type SG 1010A or Spectronics Type SE 5455-4 Clairex Type CLED-1
2. IR Photo detectors Vactec VTS4085

Circuit Notes

The LED predriver output pulses an external transistor which in turn, switches on the infrared light emitting diode at a very low duty cycle. The desired IR LED pulse period is determined by the value of the external timing resistor. The Smoke Sensitivity is adjustable through a trimmer resistor which varies the IR LED pulse width. The light sensing element is a silicon photovoltaic cell which is held at near zero bias to minimize leakage currents. The circuit can detect signals as low as 1 mV and generate an alarm. The IR LED pulse repetition rate increases when smoke is detected.

1.9 V BATTERY OPERATED IONIZATION TYPE SMOKE DETECTOR

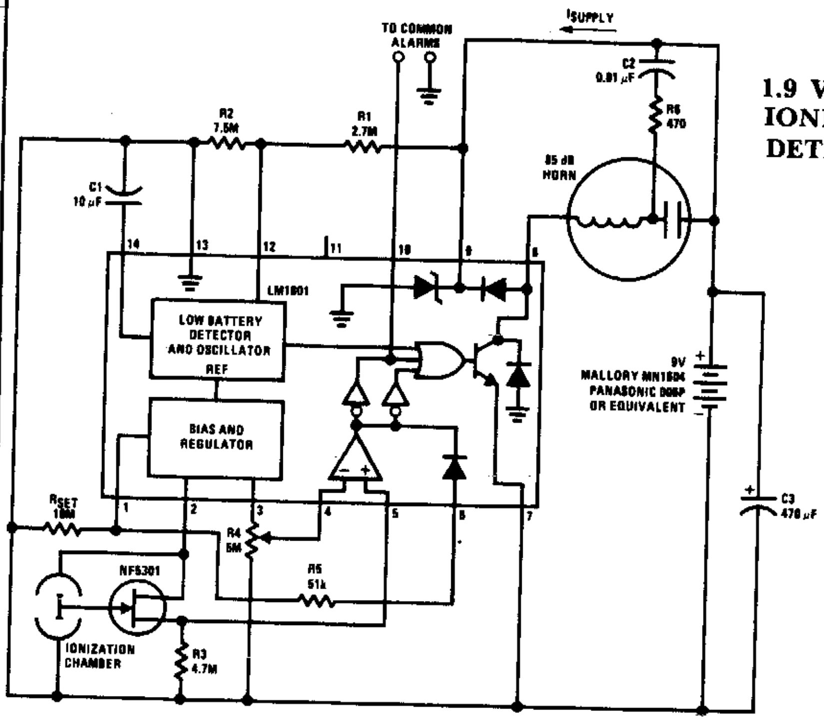


Fig. 79-2

LINE-OPERATED PHOTO ELECTRIC SMOKE ALARM USING LIGHT SENSITIVE RESISTOR (INCLUDES DETECTION OF OPEN-CIRCUITED LED)

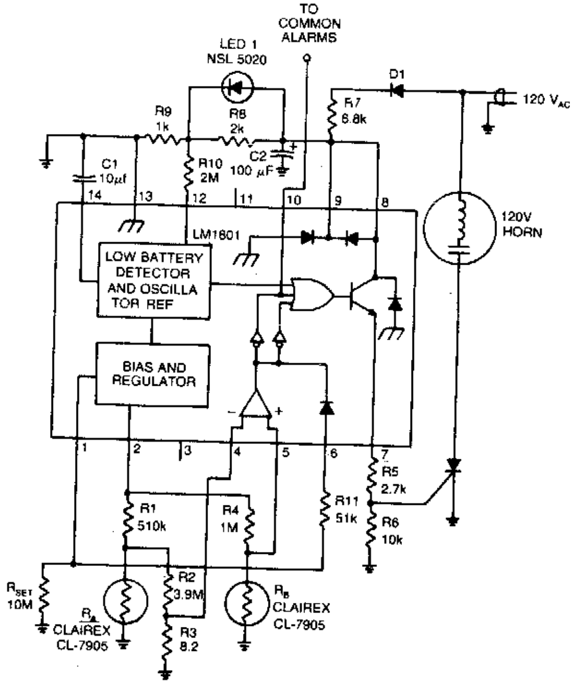


Fig. 79-3

1.9 V BATTERY OPERATED IONIZATION TYPE SMOKE DETECTOR

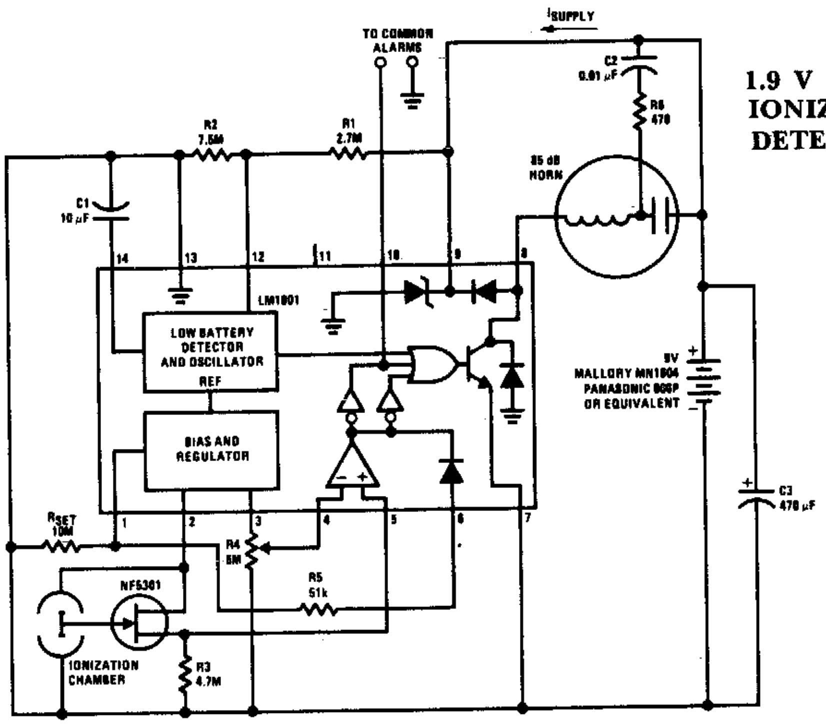


Fig. 79-2

LINE-OPERATED PHOTO ELECTRIC SMOKE ALARM USING LIGHT SENSITIVE RESISTOR (INCLUDES DETECTION OF OPEN-CIRCUITED LED)

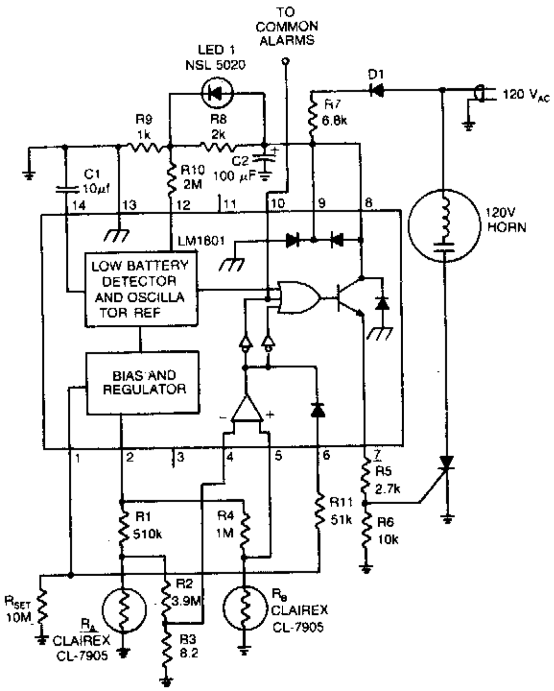


Fig. 79-3

80

Sound Effect Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VOLTAGE-CONTROLLED AMPLIFIER OR TREMOLO CIRCUIT

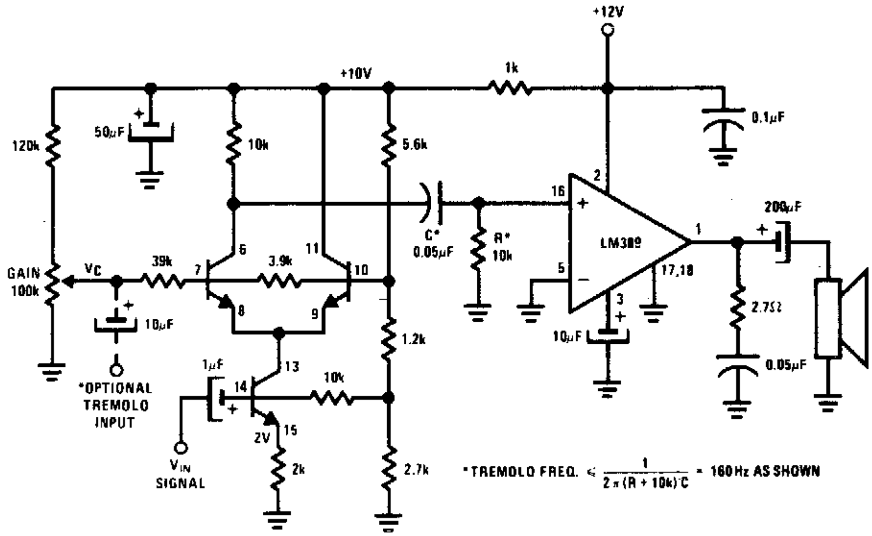


Fig. 80-1

Circuit Notes

The transistors form a differential pair with an active current-source tail. This configuration, known technically as a variable-transconductance multiplier, has an output proportional to the product of the two input signals. Multiplication occurs due to the dependence of the transistor transconductance on the emitter current bias. Tremolo (amplitude modulation of an audio frequency by a sub-audio oscillator—normally 5-15 Hz) applications require feeding the low frequency oscillator signal into the optional input shown. The gain control pot maybe set for optimum depth.

MUSIC SYNTHESIZER

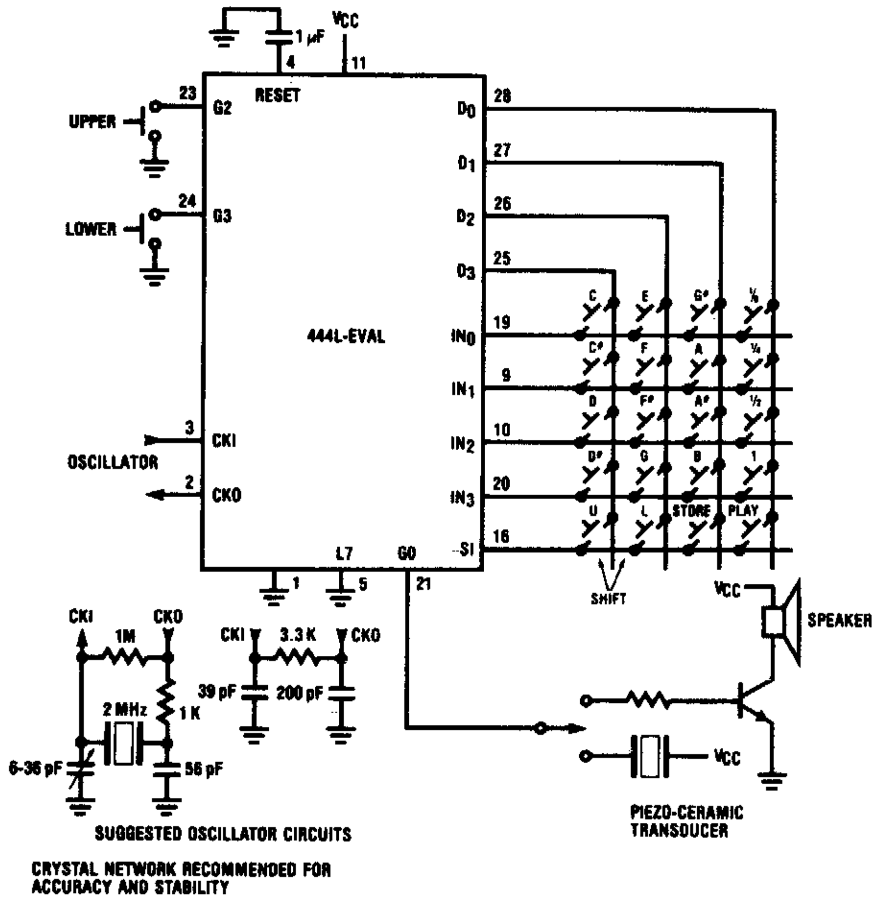


Fig. 80-2

Circuit Notes

Three modes of operation are available in the music synthesizer mode: play a note, play one of four stored tunes, or record a tune for subsequent replay.

PREPROGRAMMED SINGLE-CHIP MICROCONTROLLER FOR MUSICAL ORGAN

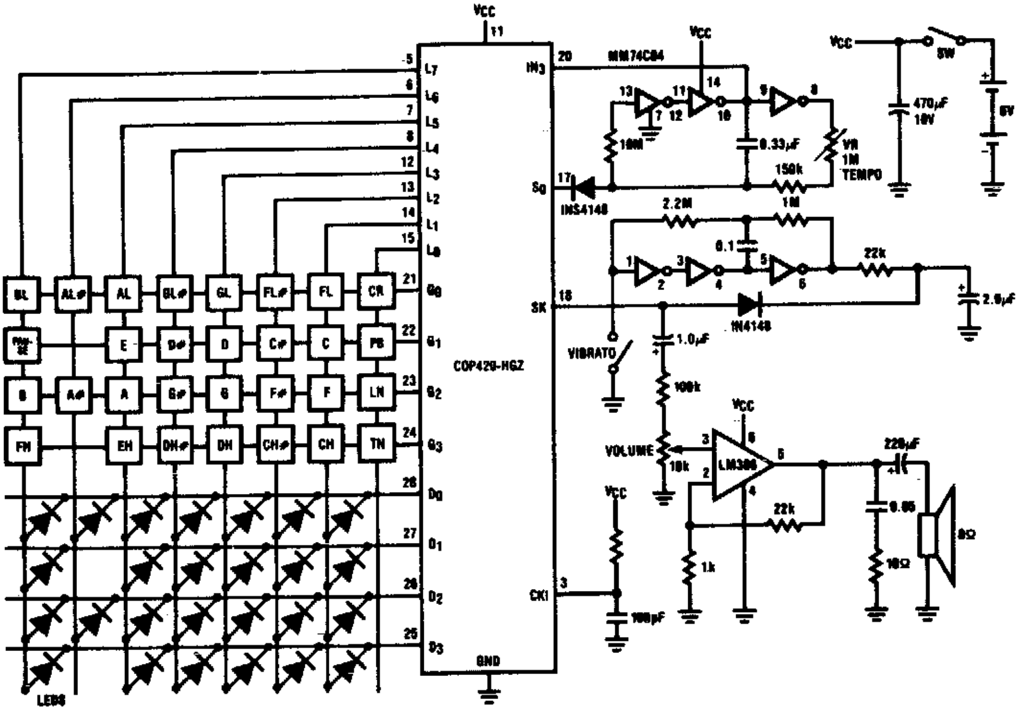


Fig. 80-3

Circuit Notes

Twenty-five musical keys and 25 LEDs are provided to denote F to F″ with half notes in between. Memory can store a played tune. There are ten preprogrammed tunes (each has an average of 55 notes) masked in the chip. Any tune can be recalled by depressing the Tune Button followed by the corresponding Sharp Key. In learn mode, the player can learn the ten preprogrammed tunes.

MUSICAL ENVELOPE GENERATOR AND MODULATOR

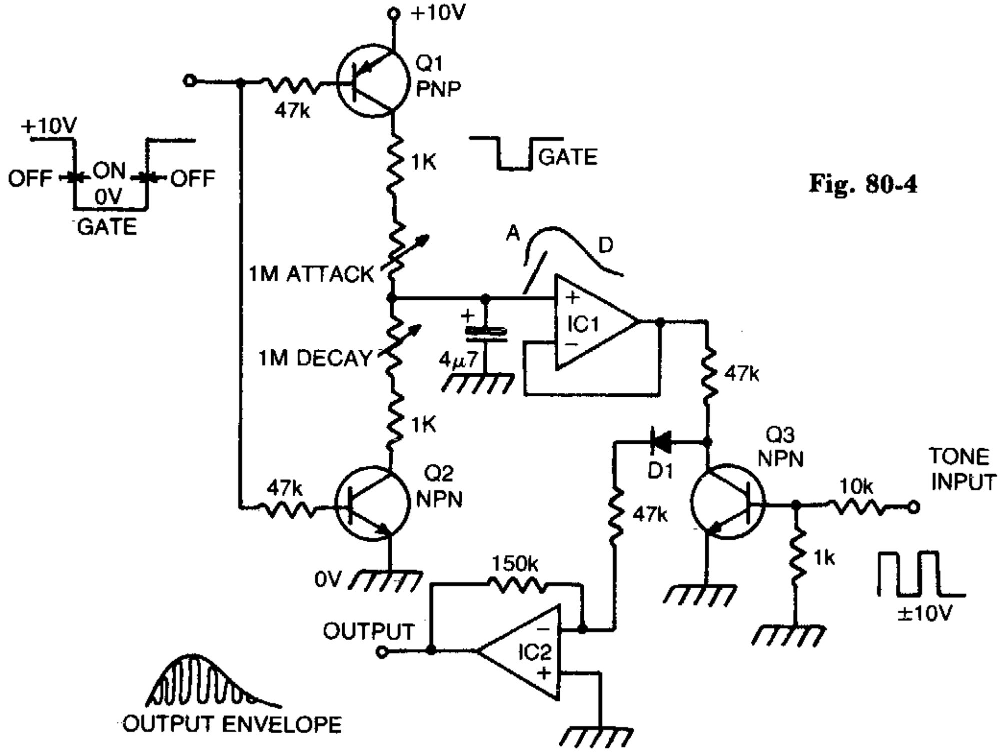


Fig. 80-4

Circuit Notes

When a gate voltage is applied, Q1 is turned on and capacitor C is charged via the attack pot in series with the 1 K resistor varying this pot, attact time constant. A fast attack gives a percussive sound, a slow attack the affect of "backward" sounds. When the gate voltage returns to its off state, Q2 is turned on and capacitor is discharged via decay pot to ground. The envelope is buffered by IC1 and applied to Q3, which is used as a transistor chopper. A musical tone in the form of a squarewave is connected to the base of Q3. This turns the transistor on or off and thus the envelope is chopped up at regular intervals, the intervals being determined by the pitch of the squarewave. The resultant waveform has the amplitude of the envelope and the harmonic structure of the squarewave. IC2 buffers the signal and D1 ensures that the envelope dies away at the end of a note.

STEREO REVERB SYSTEM

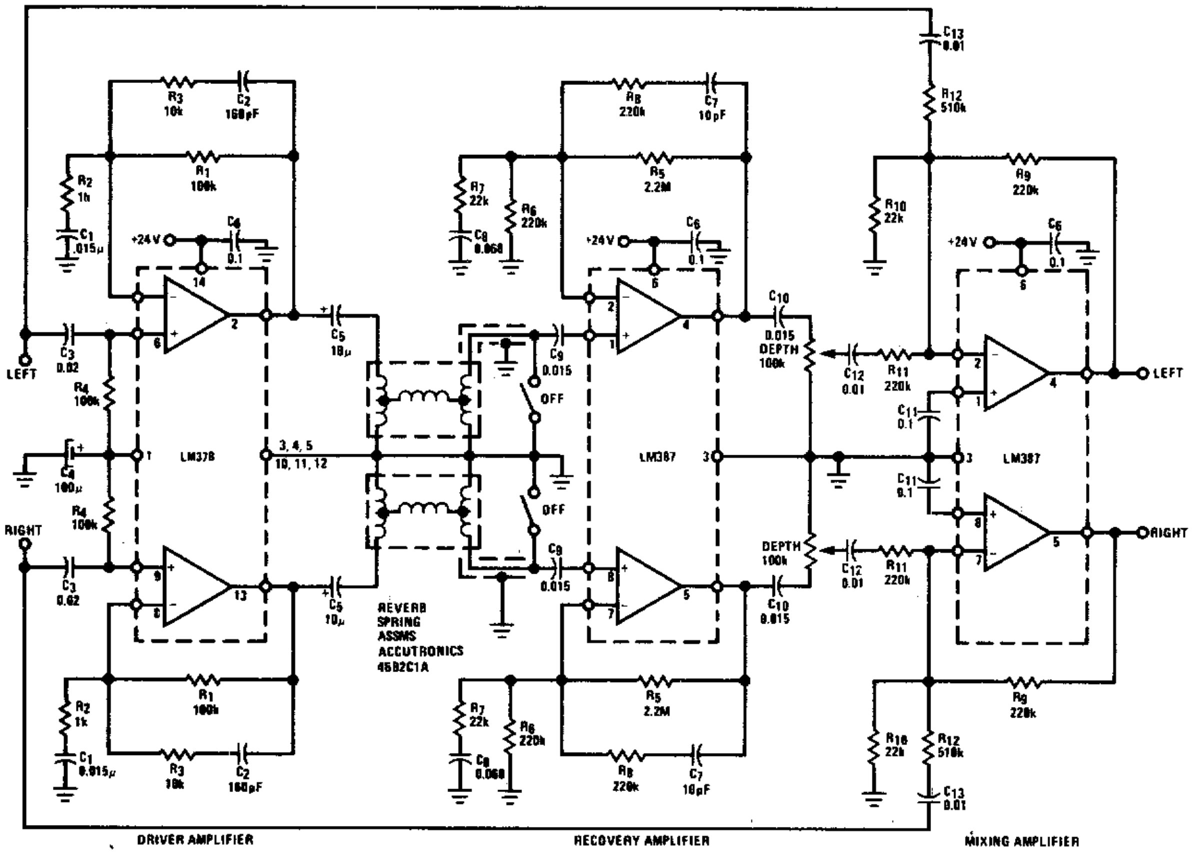


Fig. 80-5

Circuit Notes

The LM378 dual power amplifier is used as the spring driver. The recovery amplifier is a low noise dual preamplifier. Mixing of the delayed signal with the original is done with another LM387 used in an inverting summing configuration.

FOUR CHANNEL SYNTHESIZER

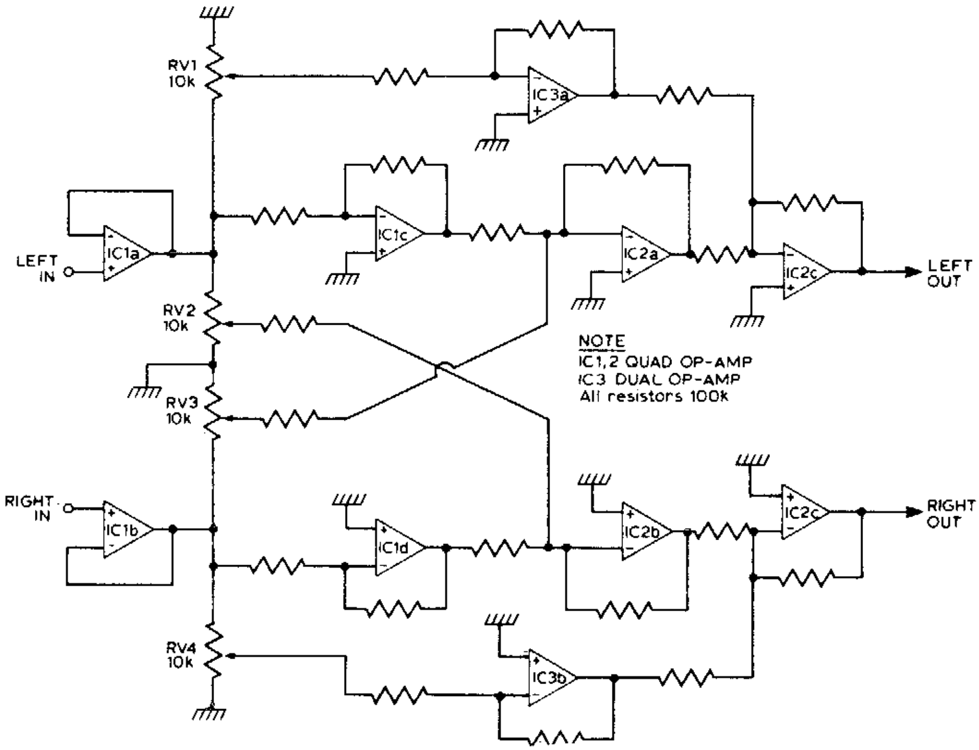


Fig. 80-6

Circuit Notes

This circuit will synthesize two rear channels for quadraphonic sound when fed with a stereo signal. The rear output for the left channel, is a combination of the left channel input 180 out of phase, added to a proportion of the right hand channel (also out of phase). The right hand rear output is obtained in a similar way.

TONE BURST GENERATOR

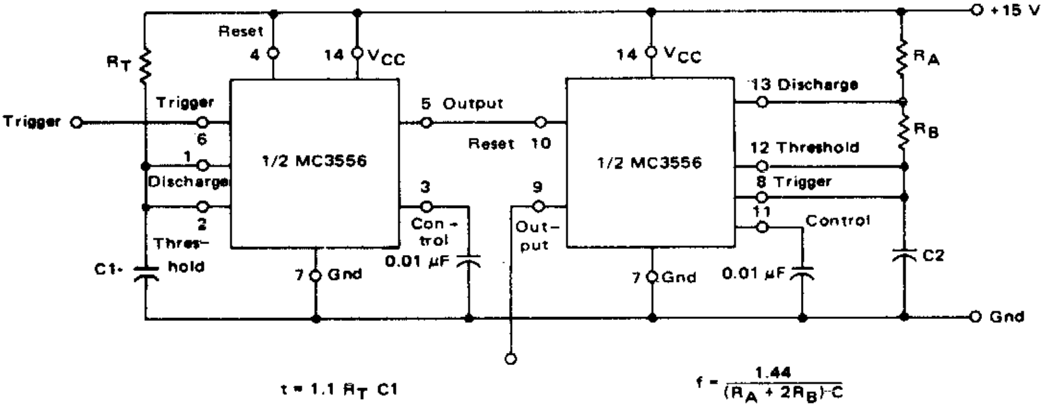


Fig. 80-7

Circuit Notes

The first timer is used as a monostable and determines the tone duration when triggered by a positive pulse at pin 6. The second timer is enabled by the high output of the monostable. It is connected as an astable and determines the frequency of the tone.

MUSICAL CHIME GENERATOR

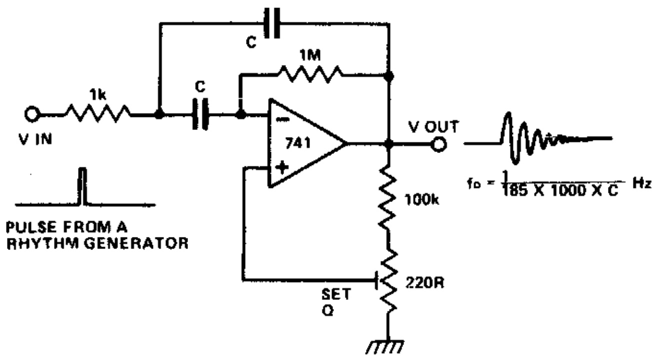


Fig. 80-8

Circuit Notes

The circuit is that of a multiple feedback bandpass filter. A short click (pulse), makes it ring with a frequency which is its natural resonance frequency. Oscillations die away exponentially and closely resemble many naturally occuring percussive or plucked sounds. The higher the Q the longer the decay time constant. High frequency resonances resemble chimes, lower frequencies sound like claves or bongos. Several circuits, all with different tuning, driven by pulses from a rhythm generator can produce an interesting pattern of sounds.

SOUND EFFECT GENERATOR

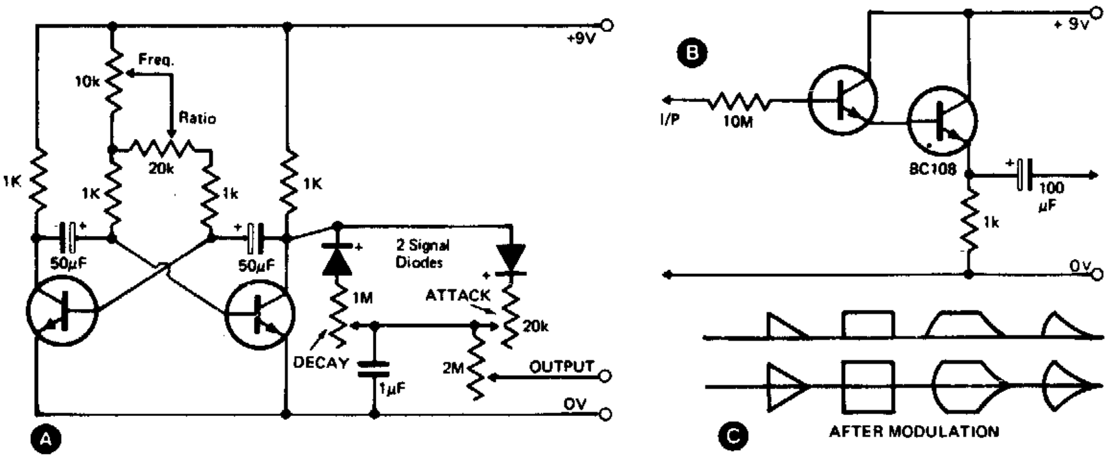


Fig. 80-9

Circuit Notes

This waveshape generator is basically a slow running oscillator with variable attack and decay. A variable amplitude (high impedance) output is available via the 2 M potentiometer. B shows an add-on circuit which should be used if a low impedance output is required. Some of the output waveforms that can be produced are shown in C.

PROGRAMMABLE BIRD SOUNDS

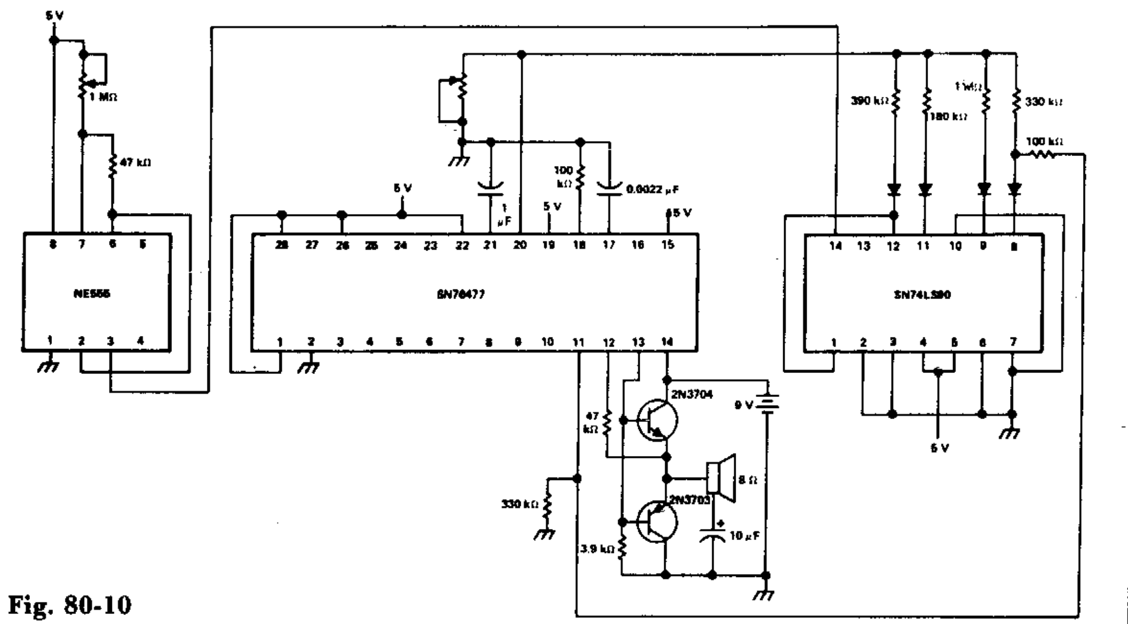


Fig. 80-10

STEREO REVERB ENHANCEMENT SYSTEM

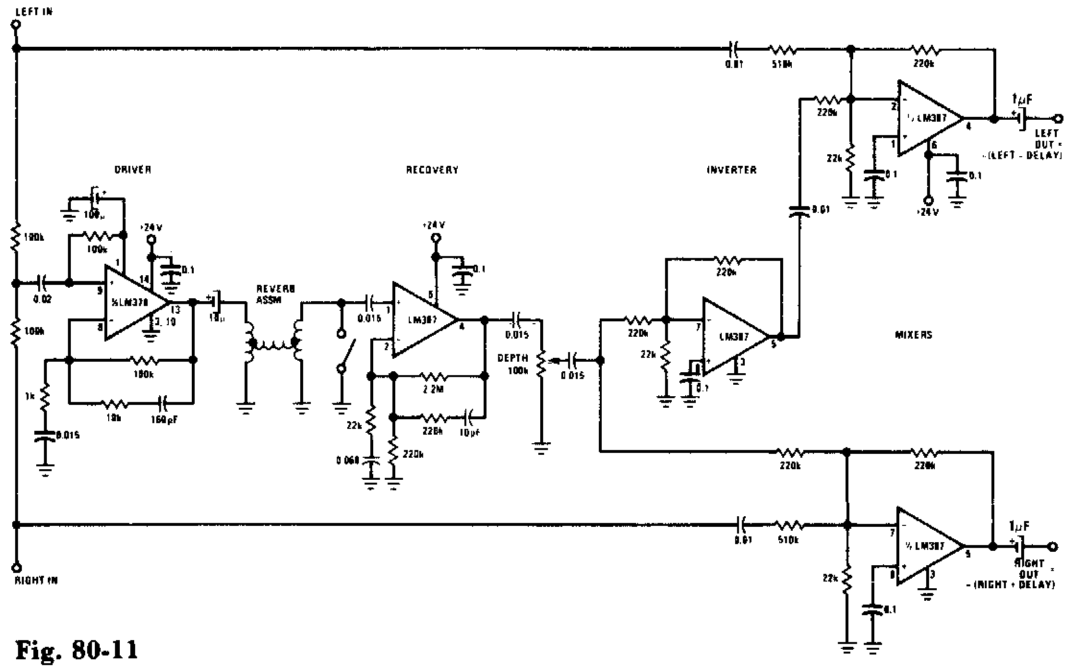


Fig. 80-11

Circuit Notes

The system can be used to synthesize a stereo effect from a monaural source such as AM radio or FM-mono broadcast, or it can be added to an existing stereo (or quad) system where it produces an exciting "opening up" special effect that is truly impressive.

SIREN/SPACE WAR/PHASOR GUN

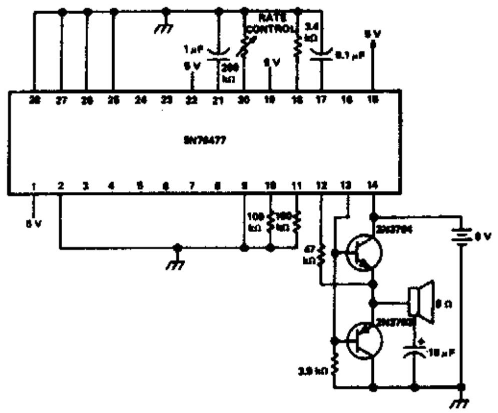


Fig. 80-12

Circuit Notes

The one shot and decay functions could be added to make an ideal phasor gun sound.

81

Sound (Audio) Operated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VOICE ACTIVATED SWITCH AND AMPLIFIER

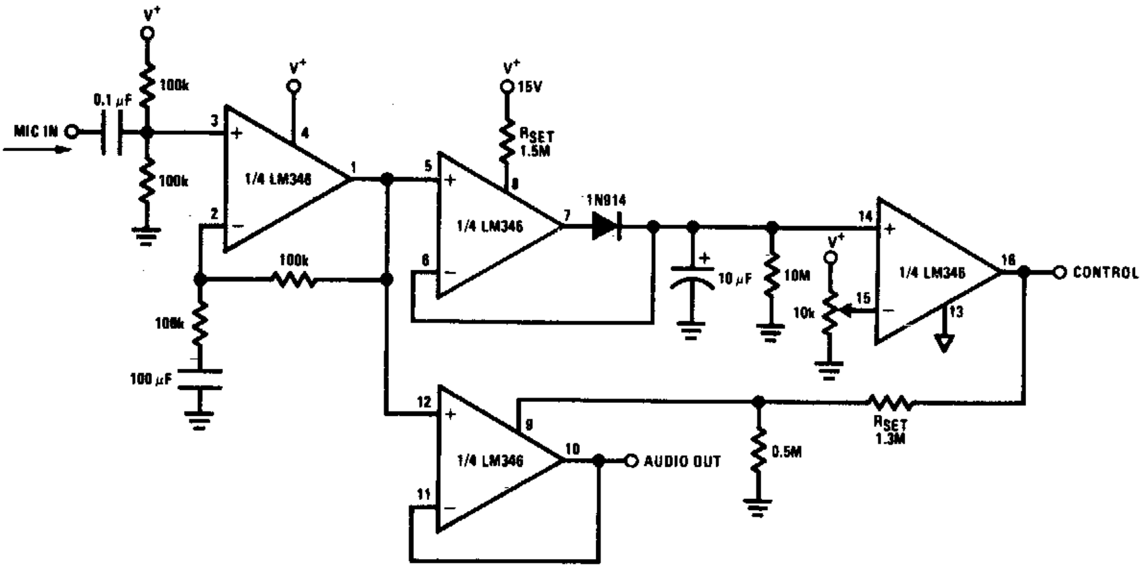


Fig. 81-1

AUDIO OPERATED RELAY

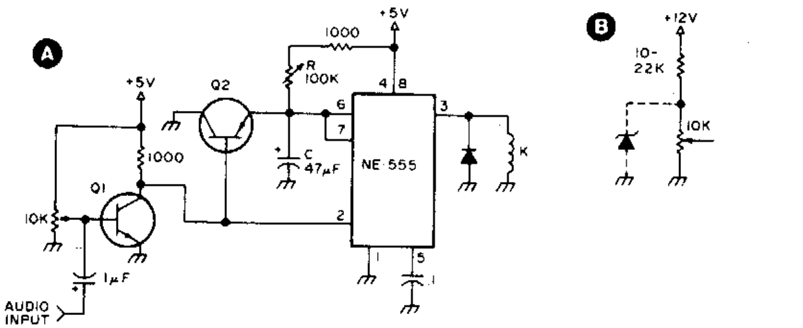
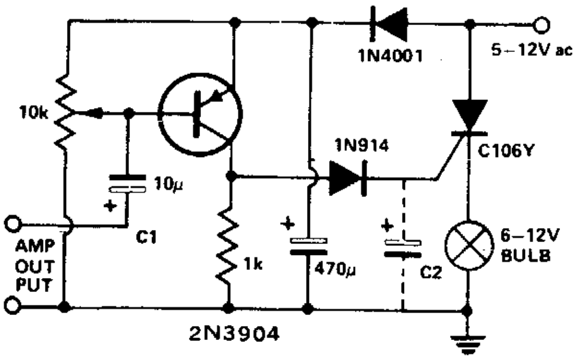


Fig. 81-2

Circuit Notes

Q1 and Q2 are general purpose transistors. The 10 K input pot is adjusted to a point just short of where Q1 turns on as indicated by K pulling in. K is any 5 V reed relay. With the values shown for R (100 K) and C (47 μF), timing values from .05 to slightly over 5 seconds can be achieved. B shows the addition of a 22 K series resistor to the 10 K input pot if a 12 V supply is used. A suitable 12 V reed relay must be used at K.

SOUND-MODULATED LIGHT SOURCE



Circuit Notes

This circuit modulates a light beam with voice or music from the output of an amplifier. If the 10 K pot is adjusted to slightly less than the V_{be} of the transistor, the circuit forms a peak detector. This drives the gate of the SCR, lighting the bulb whose brightness will vary as the sound level varies. C2 may be removed for a faster response.

Fig. 81-3

AUDIO-CONTROLLED LAMP

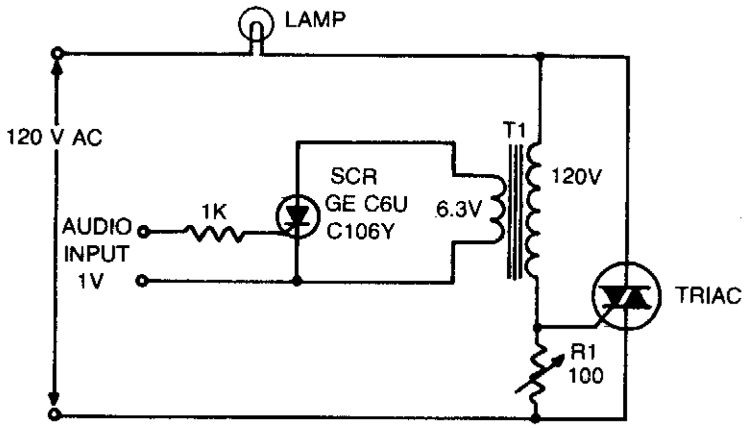


Fig. 81-4

NOTE: T1 IS A 6.3V, 1A. "FILAMENT" TRANSFORMER. ADJUST R1 FOR MAXIMUM RESISTANCE THAT WILL NOT TURN ON LAMP WITH ZERO INPUT.

Circuit Notes

This is an on-off control with isolated, low voltage input. Since the switching action is very rapid, compared with the response time of the lamp and the response of the eye, the effect produced with audio input is similar to a proportional control circuit. If the input signal to the SCR consists of phase-controlled pulses, full wave control of the lamp load is obtained.

SOUND ACTIVATED RELAY

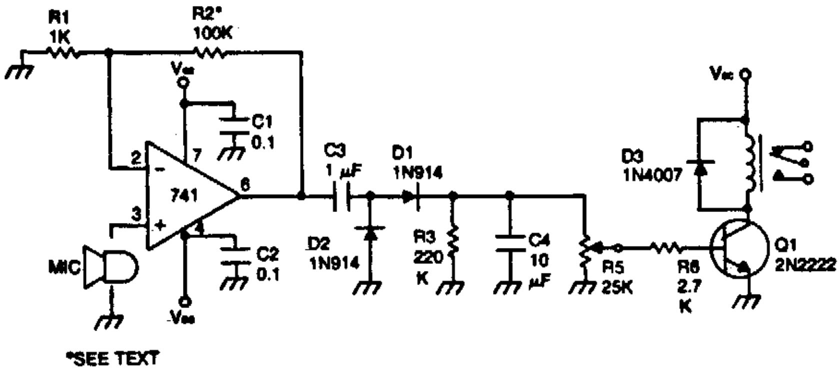


Fig. 81-5

Circuit Notes

The device remains dormat (in an off condition) until some sound causes it to turn on. The input stage is a 741 operational amplifier connected as a noninverting follower audio amplifier. Gain is approximately 100. To increase gain raise the value of R2. The amplified signal is rectified and filtered to a dc level by R4. Then R5 is set to the audio level desired to activate the relay.

SOUND OPERATED TWO-WAY SWITCH

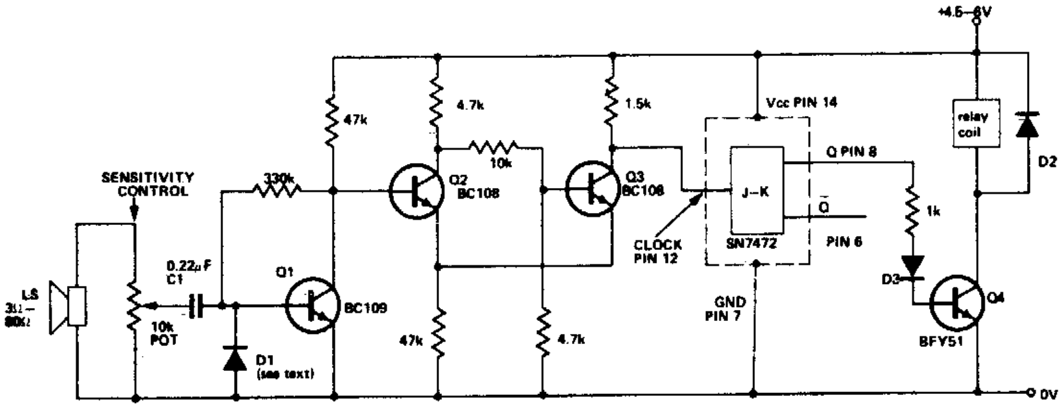


Fig. 81-6

Circuit Notes

This circuit operates a relay each time a sound of sufficient intensity is made, thus one clap of the hands will switch it one way, a second clap will revert the circuit to the original condition. Q2 and Q3 form a Schmitt trigger. The JK flip-flop is used as a bistable whose output changes state every time a pulse is applied to the clock input (pin 12). Q4 allows the output to drive a relay.

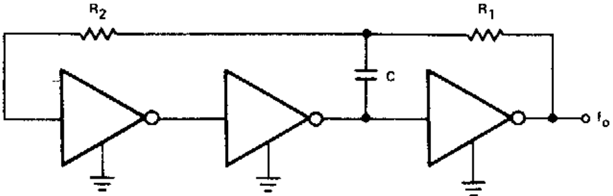
82

Square Wave Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

R/C OSCILLATOR

$$f_O \approx \frac{1}{2\,C[0.41\,R_P + 0.70\,R_1]}, \quad R_P = \frac{R_1\,R_2}{R_1 + R_2}$$



Gates are 74C04

a. If $R_1 = R_2 = R_1$, $f \cong 0.55/RC$

b. If $R_2 >> R_1$, $f \cong 0.45/R_1C$

c. If $R_2 << R_1$, $f \cong 0.72/R_1C$

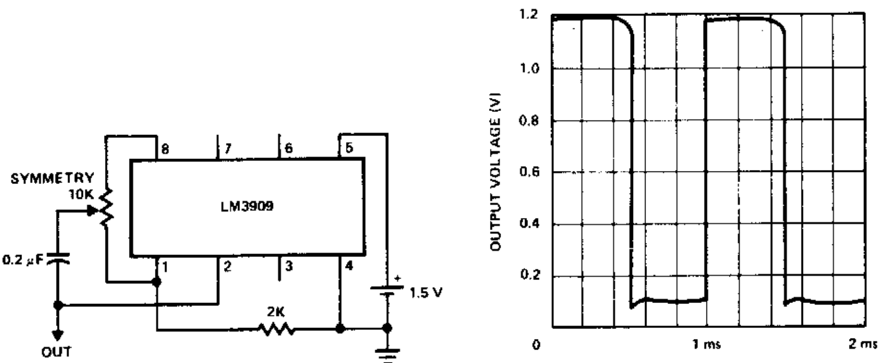
a. $f = 120$ kHz, $C = 420$ pF
   $R_1 = R_2 \approx 10.9\ k\,\Omega$

b. $f = 120$ kHz, $C = 420$ pF, $R_2 = 50\ k\,\Omega$
   $R_1 = 8.93\ k\,\Omega$

c. $f = 120$ kHz, $C = 220$ pF, $R_2 = 5\ k\,\Omega$
   $R_1 = 27.3\ k\,\Omega$

Fig. 82-1

1 kHz SQUARE WAVE OSCILLATOR



Note: Output Voltage Through a 10K Load to Ground

Fig. 82-2

TTL OSCILLATOR

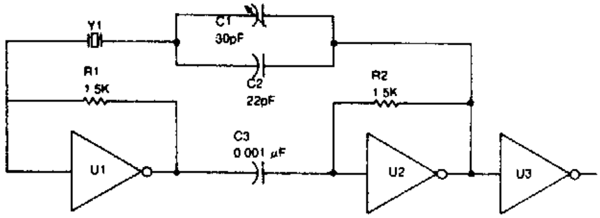


Fig. 82-3

Circuit Notes

TTL inverter stages, U1 and U2, are cross-connected with a crystal Y1. A resistor in each stage biases the normally digital gates into a region where they operate as amplifiers. Inverter stage U3 is used as a buffer.

SQUARE WAVE OSCILLATOR

Oscillator Frequency for Various Capacitor Values

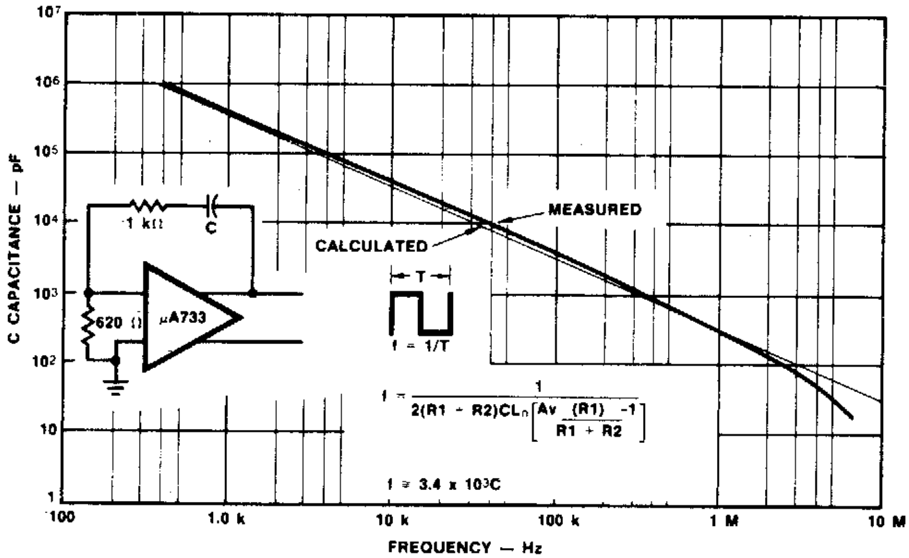


Fig. 82-4

ADJUSTABLE TTL CLOCK (MAINTAINS 50% DUTY CYCLE)

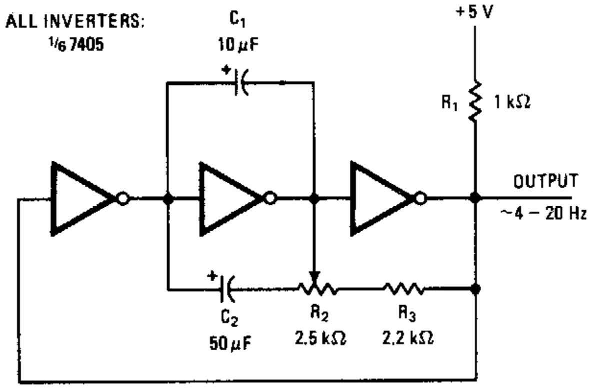


Fig. 82-5

Circuit Notes

Symmetry of the square-wave output is maintained by connecting the right side of R2 through resistor R3 to the output of the third amplifier stage. This changes the charging current to the capacitors in proportion to the setting of frequency-adjusting potentiometer R2. Thus, a duty cycle of 50% is constant over the entire range of oscillation. The lower frequency limit is set by capacitor C2. With the components shown, the frequency of oscillation can be varied by R2 from about 4 to 20 hertz. Other frequency ranges can be obtained by changing the values of C1 and R3, which control the upper limit of oscillation, or C2, which limits the low-frequency end.

SQUARE WAVE OSCILLATOR

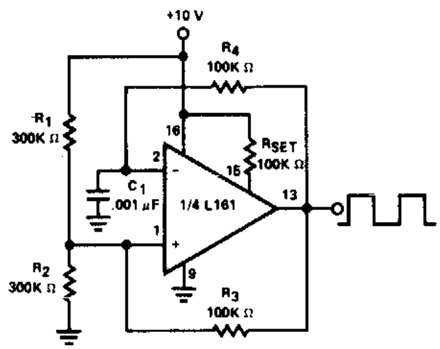
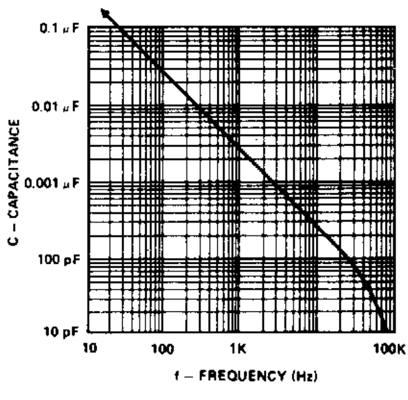


Fig. 82-6



Frequency vs the Value of C_1 for the Squarewave Oscillator

Circuit Notes

This generator is operable to over 100 kHz. The low frequency limit is determined by C1. Frequency is constant for supply voltages down to +5 V.

OSCILLATOR/CLOCK GENERATOR

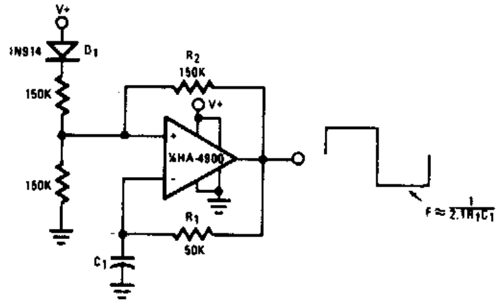
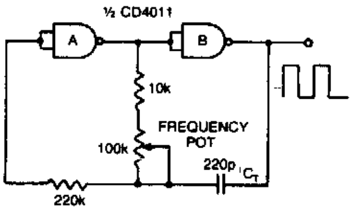


Fig. 82-7

Circuit Notes

This self-starting fixed frequency oscillator circuit gives excellent frequency stability. R1 and C1 comprise the frequency determining network while R2 provides the regenerative feedback. Diode D1 enhances the stability by compensating for the difference between V_{OH} and V_{Supply}. In applications where a precision clock generator up to 100 kHz is required, such as in automatic test equipment, C1 may be replaced by a crystal.

CMOS OSCILLATOR



Circuit Notes

Varying the 100 K pot changes the discharge rate of C_T and hence the frequency. A square wave output is generated. The maximum frequency using CMOS is limited to 2 MHz.

Fig. 82-8

FREE-RUNNING SQUARE-WAVE OSCILLATOR

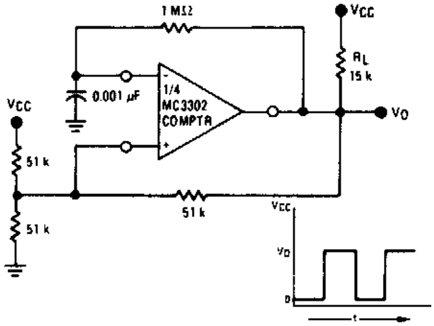


Fig. 82-9

PRECISION SQUARER

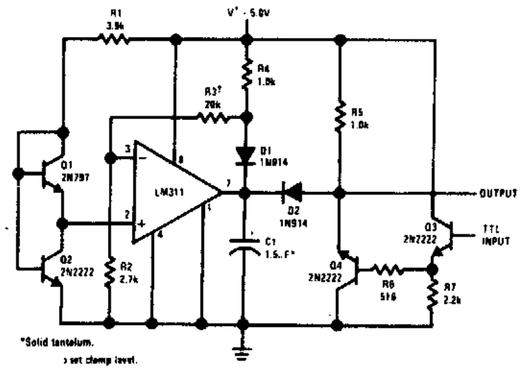
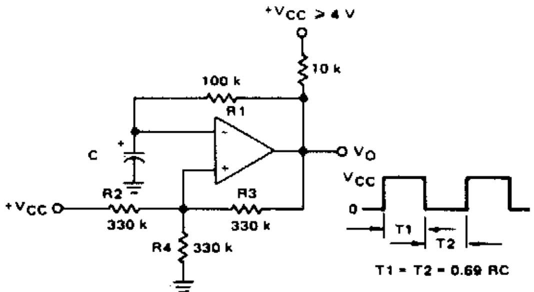


Fig. 82-10

SQUARE WAVE OSCILLATOR



$T1 = T2 = 0.69\ RC$

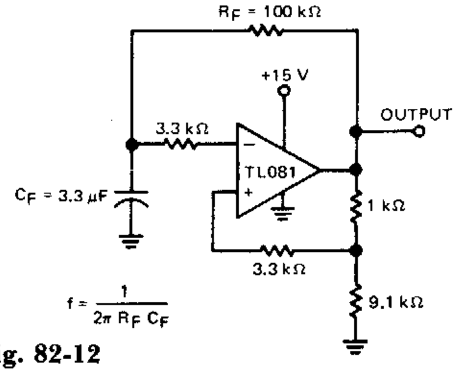
$f \approx \frac{7.2}{C(\mu F)}$

$R2 = R3 = R4$

$R1 \approx R2//R3//R4$

Fig. 82-11

0.5 Hz SQUARE-WAVE OSCILLATOR



$f = \frac{1}{2\pi R_F C_F}$

Fig. 82-12

SIMPLE TRIANGLE/SQUARE WAVE OSCILLATOR

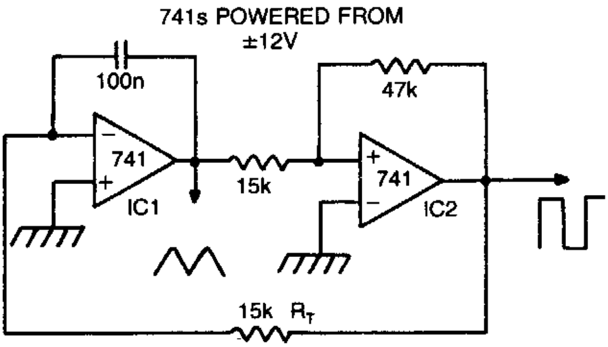


Fig. 82-13

Circuit Notes

By making R_T variable it is possible to alter the operating frequency over a 100 to 1 range. Versatile triangle/square-wave oscillator has a possible frequency range of 0.1 Hz to 100 kHz.

SQUAREWAVE OSCILLATOR

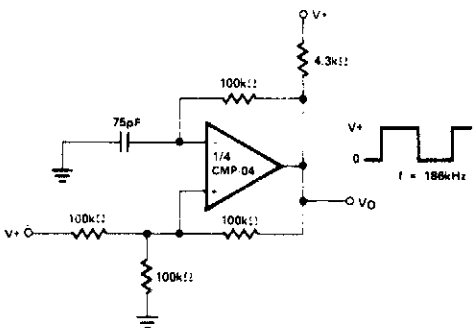


Fig. 82-14

83

Stereo Balance Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

STEREO BALANCE METER

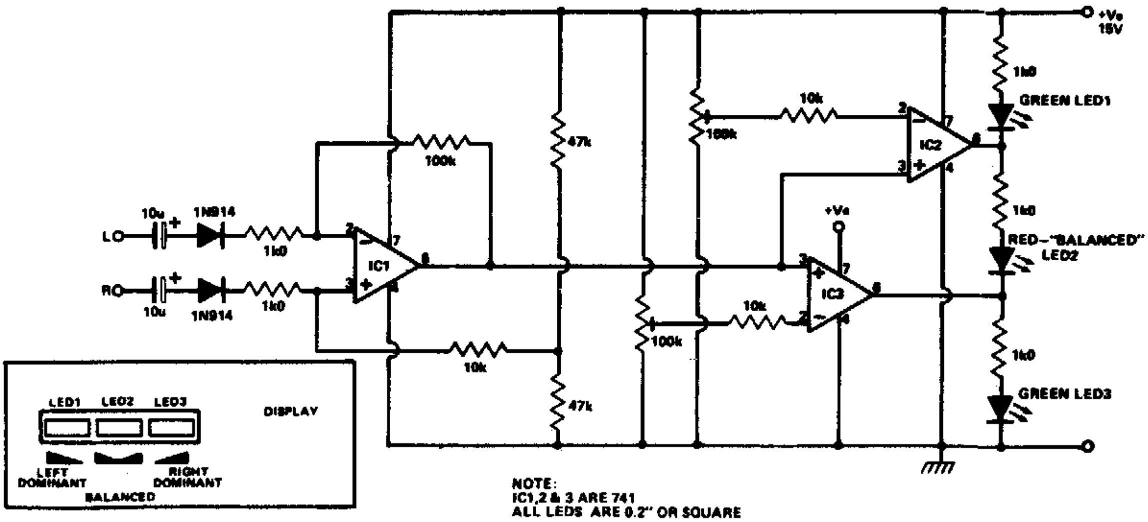


Fig. 83-1

Circuit Notes

Outputs from each channel are fed to the two inputs of IC1 connected as a differential amplifier. IC2 and 3 are driven by the output of IC1. Output of IC1 is connected to the noninverting inputs of IC2 and 3. If the output of IC1 approaches the supply rail, the outputs of ICs 2 and 3 will also go high, illuminating LED3. This would happen if the right channel were dominating. If the left channel was dominant, the outputs of ICs 2 and 3 would be low, illuminating LED1. If the two channels are equal in amplitude, the outputs of ICs 2 and 3 would be high and low respectively, lighting up LED2.

STEREO BALANCER

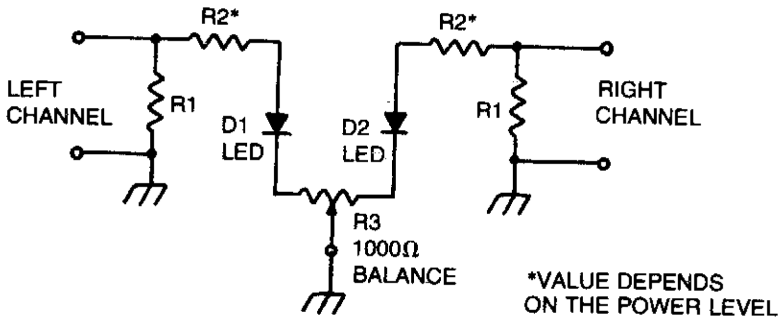


Fig. 83-2

Circuit Notes

This circuit will allow you to set the gain of two stereo channels to the same level. The signal across the two channel-load resistors is sampled by resistors R2. (Values of these resistors will depend upon the power level.) For most 20 milliampere LED, use approximately 2.5 K per watt. (For a 10-watt system use a 25,000 ohm resistor.) To set up, short the two inputs and connect them to one channel of a power amplifier. Apply a signal and adjust R3 until both LEDs glow at the same brightness level. The balancer is ready for use. Connect the inputs of the stereo balancer across the output of the power amplifier, and then turn up either the independent volume controls, or the balance control until both LEDs glow at the same level. To use this circuit in-line with loudspeakers, disconnect both R1s, and use the speakers as the load.

STEREO BALANCE METER

PARTS LIST FOR STEREO BALANCE METER

D1, D2—Silicon rectifier rated 100 PIV at any low current
M1—Zero-center DC mA meter (see text)
R1, R2—1000-ohm, ½-watt resistor, 5% or 1%

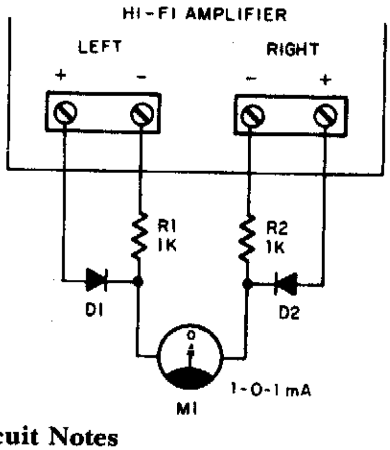


Fig. 83-3

Circuit Notes

Play any stereo disc or tape and then set the amplifier to mono. Adjust left and right channel balance until meter M1 indicates zero; then the left and right output level are identical.

84

Switches

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DTL-TTL CONTROLLED BUFFERED ANALOG SWITCH

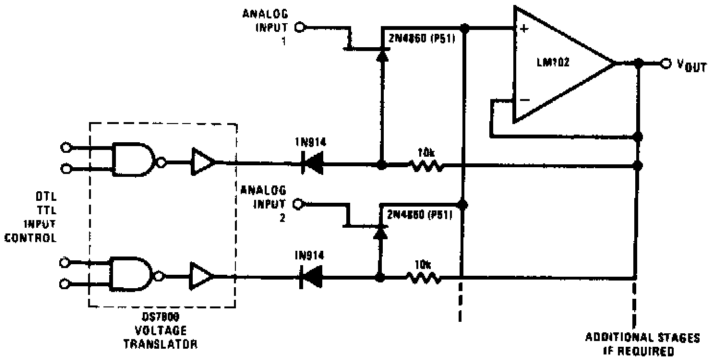


Fig. 84-1

Circuit Notes

This analog switch uses the 2N4860 JFET for its 25 ohm r_{on} and low leakage. The LM102 serves as a voltage buffer. This circuit can be adapted to a dual trace oscilloscope chopper. The DS7800 monolithic IC provides adequate switch drive controlled by DTL/TTL logic levels.

HIGH TOGGLE RATE HIGH FREQUENCY ANALOG SWITCH

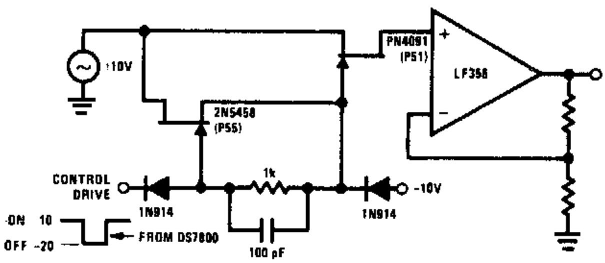


Fig. 84-2

Circuit Notes

Commutator circuit provides low impedance gate drive to the PN4091 analog switch for both on and off drive conditions. This circuit also approaches the ideal gate drive conditions for high frequency signal handling by providing a low ac impedance for off drive and high ac impedance for on drive to the PN4091

DIFFERENTIAL ANALOG SWITCH

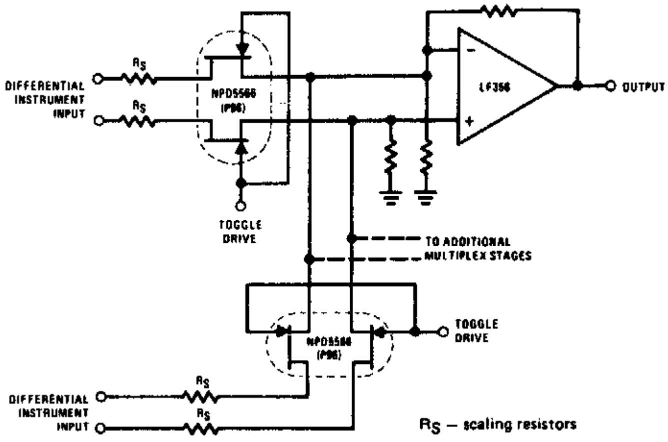


Fig. 84-3

Circuit Notes

The NPD5566 monolithic dual is used in a differential multiplex application where $R_{ds(ON)}$ should be closely matched. Since $R_{ds(ON)}$ for the monolithic dual tracks at better than ±1% over wide temperature ranges (−25° C to +125° C), this makes it an unusual but ideal choice for an accurate multiplexer. This close tracking greatly reduces errors due to common-mode signals.

HIGH FREQUENCY SWITCH

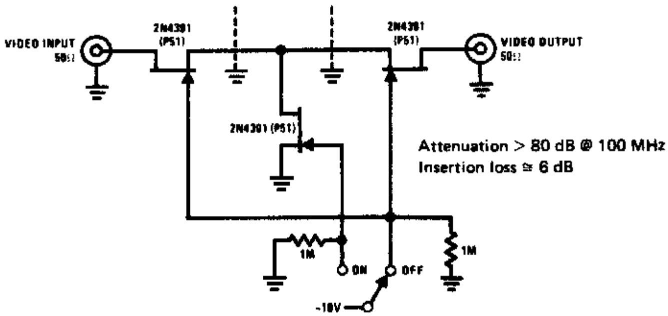
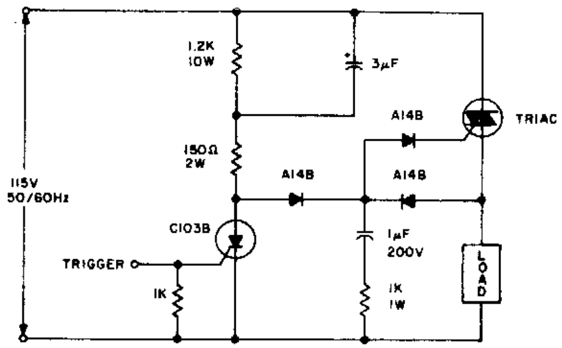


Fig. 84-4

Circuit Notes

The 2N4391 provides a low ON resistance of 30 ohm and a high OFF impedance (< 0.2 pF) when off. With proper layout and an ideal switch, the performance stated above can be readily achieved.

TRIAC ZERO VOLTAGE SWITCHING

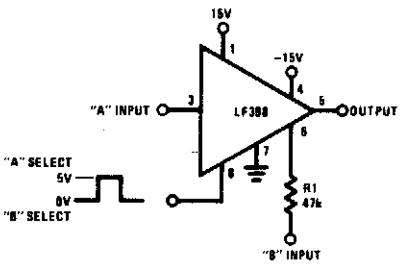


Circuit Notes

The triac will be gated on at the start of the positive half cycle by current flow through the 3 μF capacitor as long as the C103 SCR is off. The load voltage then charges up the 1 μF capacitor so that the triac will again be energized during the subsequent negative half cycle of line voltage. A selected gate triac is required because of the III+ triggering mode.

Fig. 84-5

TWO-CHANNEL SWITCH



	A	B
Gain	1 ±0.02%	1 ±0.2%
Z_{IN}	$10^{10}\Omega$	47 kΩ
BW	≃1 MHz	≈400 kHz
Crosstalk @ 1 kHz	−90 dB	−90 dB
Offset	≤ 6 mV	≤ 75 mV

Fig. 84-6

10 A, 25 Vdc SOLID STATE RELAYS

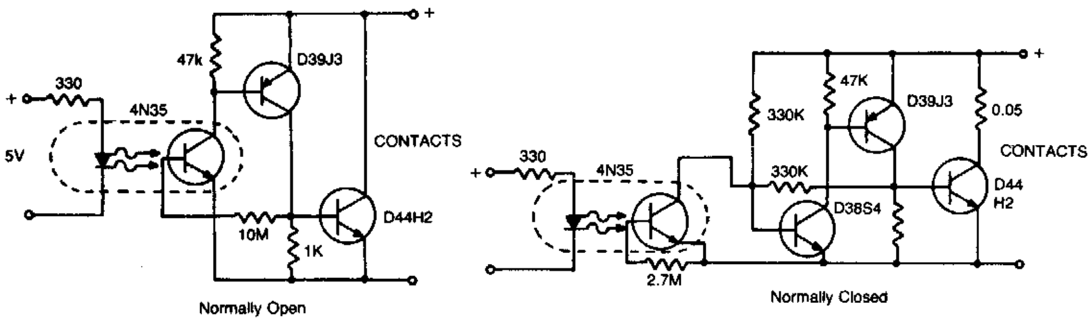


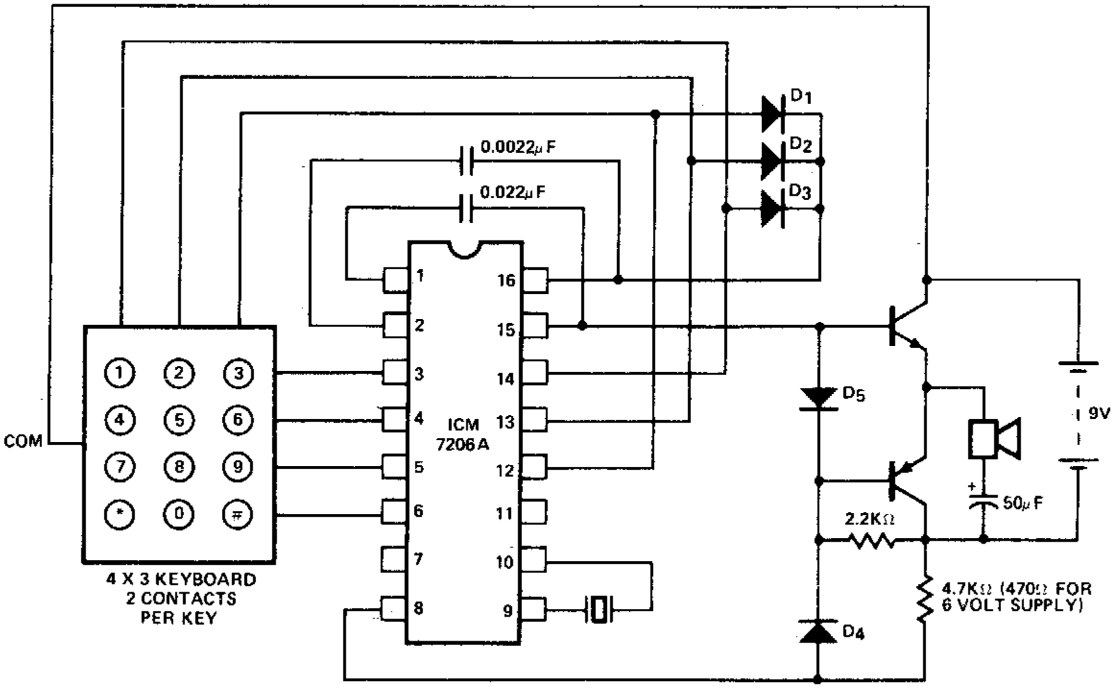
Fig. 84-7

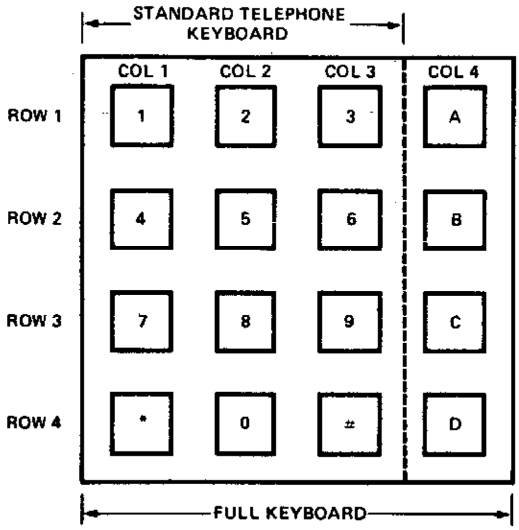
85

Telephone Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PORTABLE TONE GENERATOR





KEY	LOW BAND FREQ. Hz	HI BAND FREQ. Hz
1	697	1209
2	697	1336
3	697	1477
4	770	1209
5	770	1336
6	770	1477
7	852	1209
8	852	1336
9	852	1477
*	941	1209
0	941	1336
#	941	1477
A	697	1633
B	770	1633
C	852	1633
D	941	1633

Fig. 85-1

TELEPHONE STATUS MONITOR USING AN OPTOISOLATOR

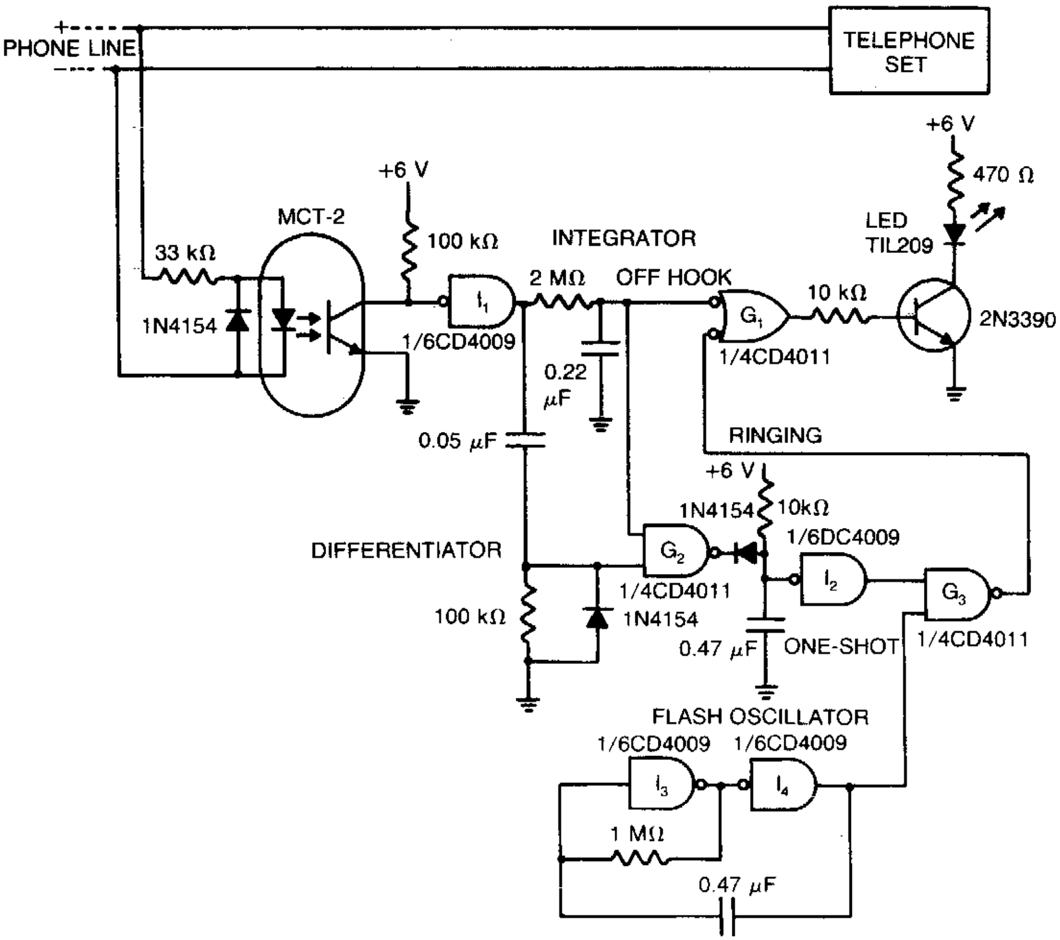
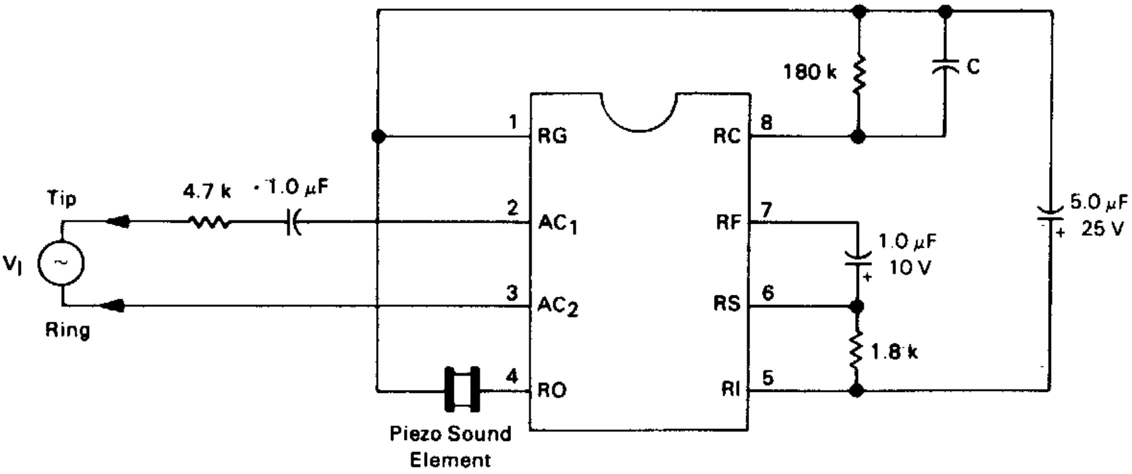


Fig. 85-2

Circuit Notes

The LED indicates the status of a remote telephone. The light is off if the phone is hung up. It shines steadily if the phone is off hook, and it flashes on and off while phone rings and for 5 seconds after ringing stops. The flashing oscillator operates continuously but can drive the LED only when a ringing signal discharges the one shot capacitor to enable NAND gate G3. Thus, one oscillator handles several phone lines.

TELEPHONE TONE RINGER



MC34012-1: C = 1000 pF
MC34012-2: C = 500 pF
MC34012-3: C = 2000 pF

Fig. 85-3

Circuit Notes

This is a complete telephone bell replacement circuit with minimum external components with on-chip diode bridge and transient protection and direct drive for piezoelectric transducers.

F.C.C. APPROVED TELEPHONE TONE RINGER

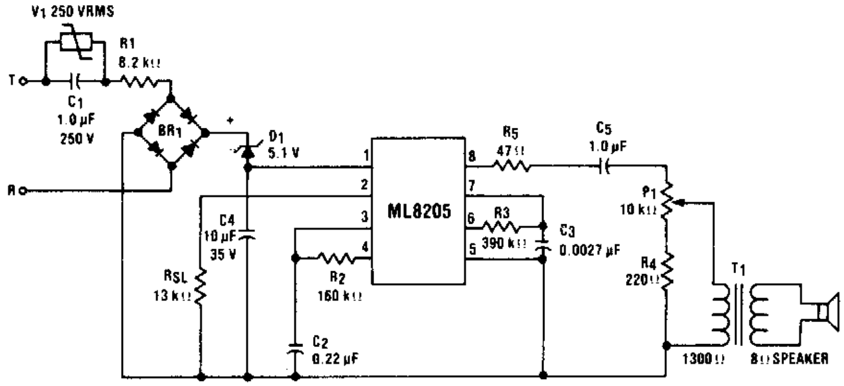


Fig. 85-4

TELEPHONE OR EXTENSION TONE RINGER

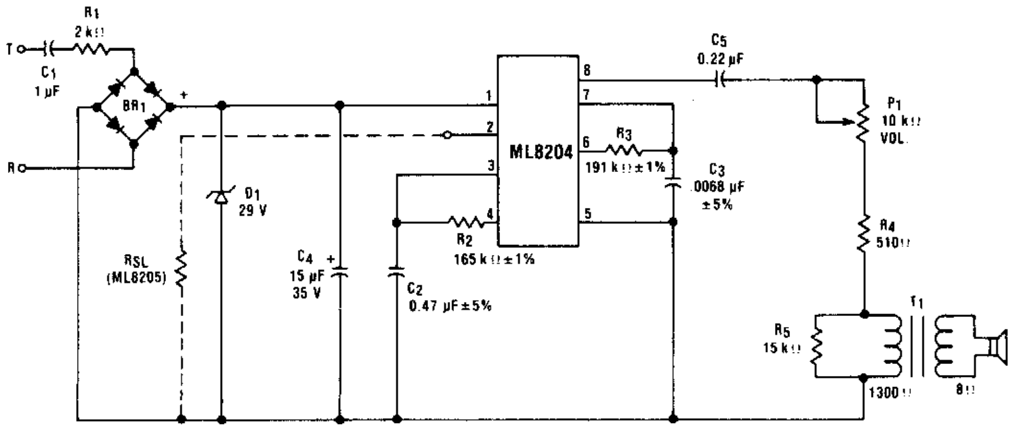


Fig. 85-5

Circuit Notes

This circuit uses ML8204/ML8205 devices. With the components shown, the output frequency chops between 512 Hz (f_{H1}) and 640 Hz (f_{H2}) at a 10 Hz (f_L) rate.

TELEPHONE LINE MONITOR

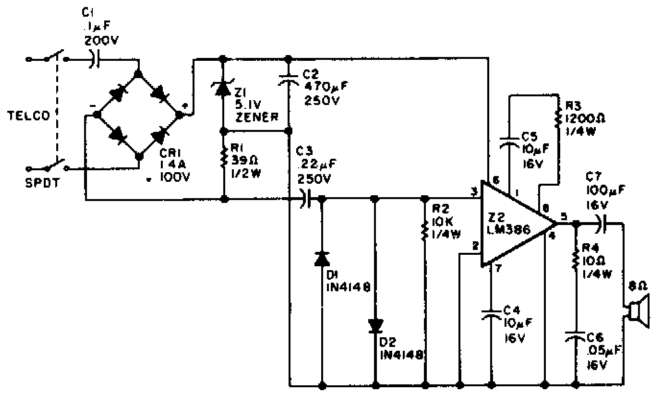


Fig. 85-6

Circuit Notes

Using rectified audio as a power supply, this monitor will send the telephone line audio into an 8 ohm speaker.

TONE DIAL GENERATOR

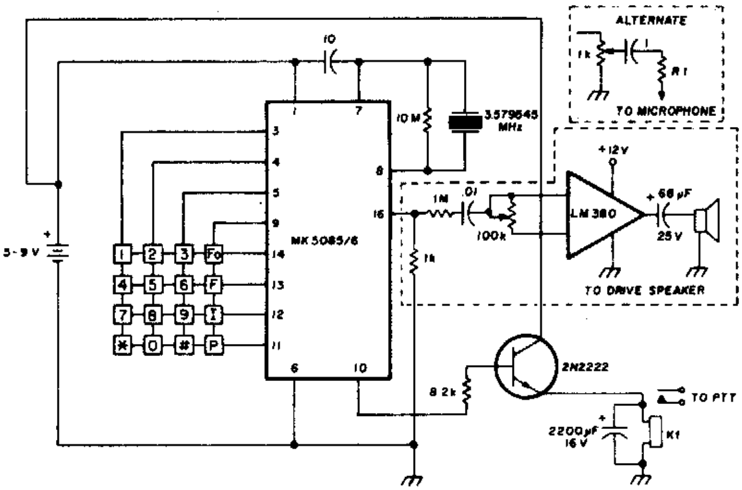


Fig. 85-7

Circuit Notes

The circuit requires a minimum of parts and uses a low cost standard 3.579545-MHz television color-burst crystal. The speaker can be eliminated and the output fed directly into the microphone input of a transmitter.

TONE DIAL ENCODER

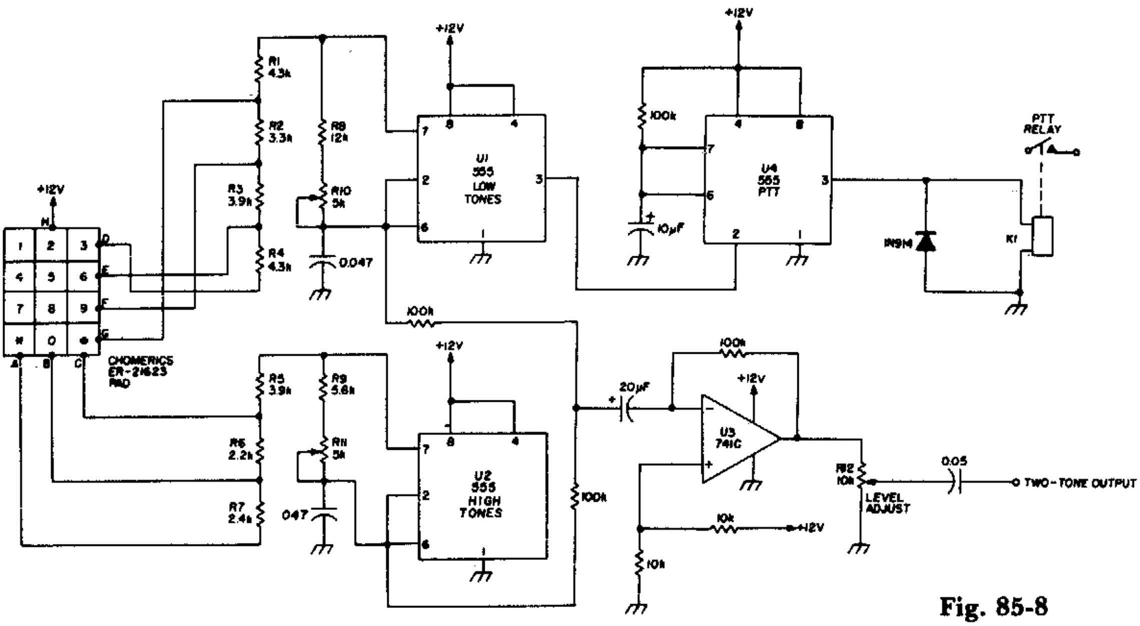
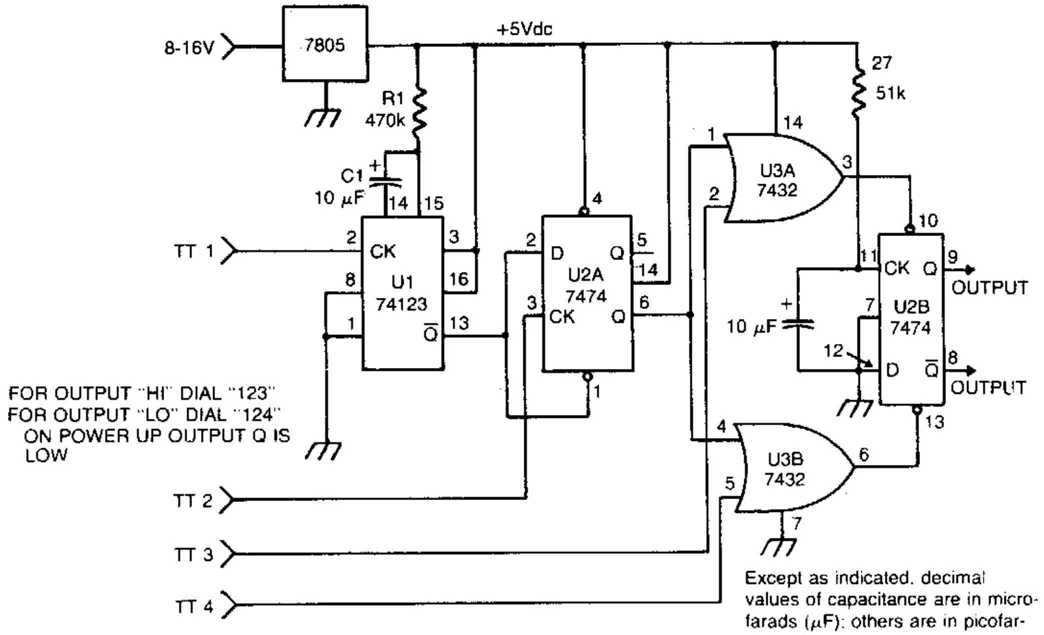


Fig. 85-8

Circuit Notes

Tone dial encoder with automatic PTT control uses the 555 timers.

TONE DIAL SEQUENCE DECODER



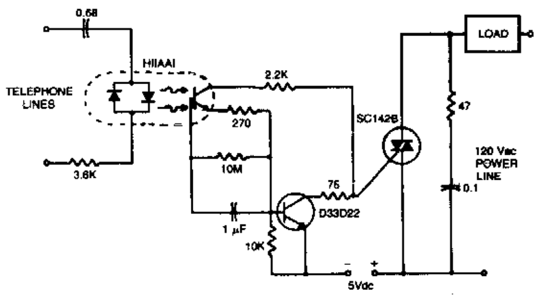
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

Fig. 85-9

Circuit Notes

The circuit takes active low inputs from a Touch Tone decoder and reacts to a proper sequence of digits. The proper sequence is determined by which Touch Tone digits the user connects to the sequence decoder inputs TT1, TT2, TT3, and TT4.

REMOTE RING EXTENDER SWITCH

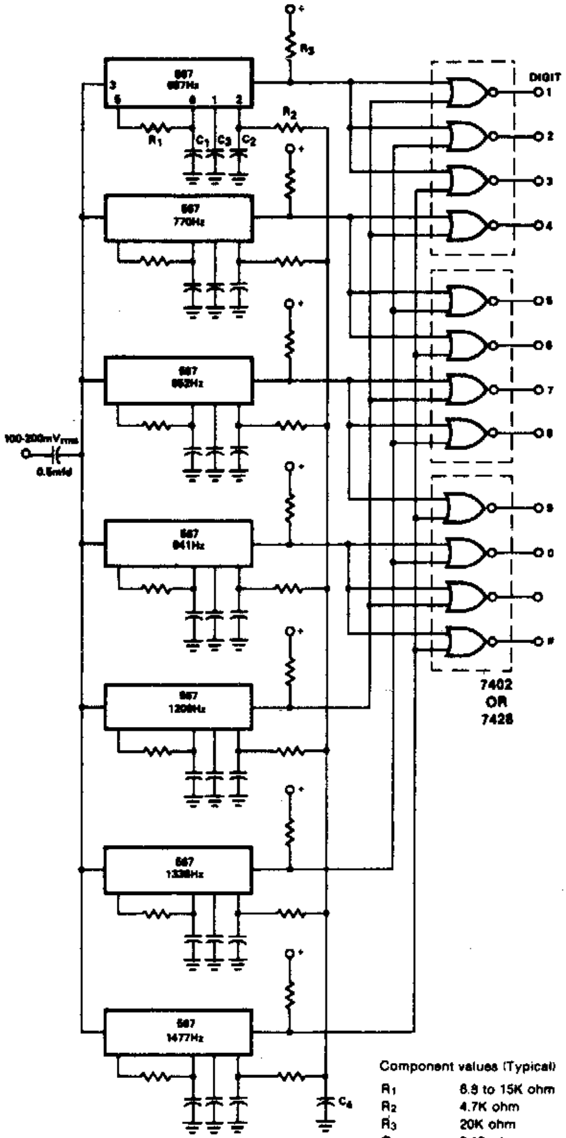


Circuit Notes

The circuit can operate lamps and buzzers from the 120 V, 60 Hz power line while maintaining positive isolation between the telephone line and the power line. Use of the isolated tab triac simplifies heat sinking by removing the constraint of isolating the triac heat sink from the chassis.

Fig. 85-10

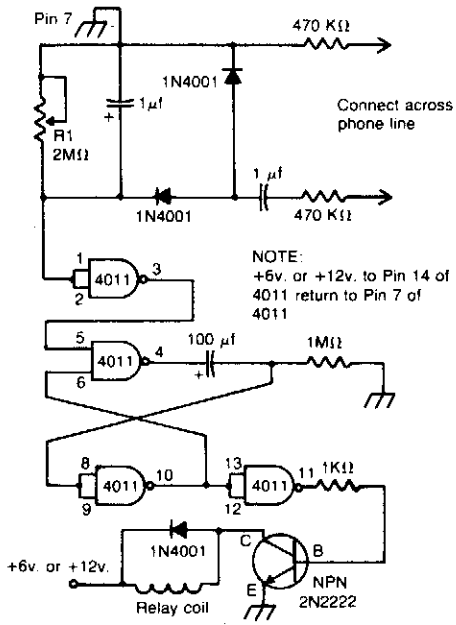
TONE DIAL DECODER



Component values (Typical)	
R_1	6.8 to 15K ohm
R_2	4.7K ohm
R_3	20K ohm
C_1	0.10mfd
C_2	1.0mfd 5V
C_3	2.2mfd 6V
C_4	250 µF 6V

Fig. 85-11

TELEPHONE RELAY



Circuit Notes

Connected across the bell circuit of phone, this circuit closes a relay when the phone is ringing. Use the delay contacts to actuate any bell, siren, buzzer or lamp.

Fig. 85-12

TELEPHONE-CONTROLLED TAPE STARTER (TCTS)

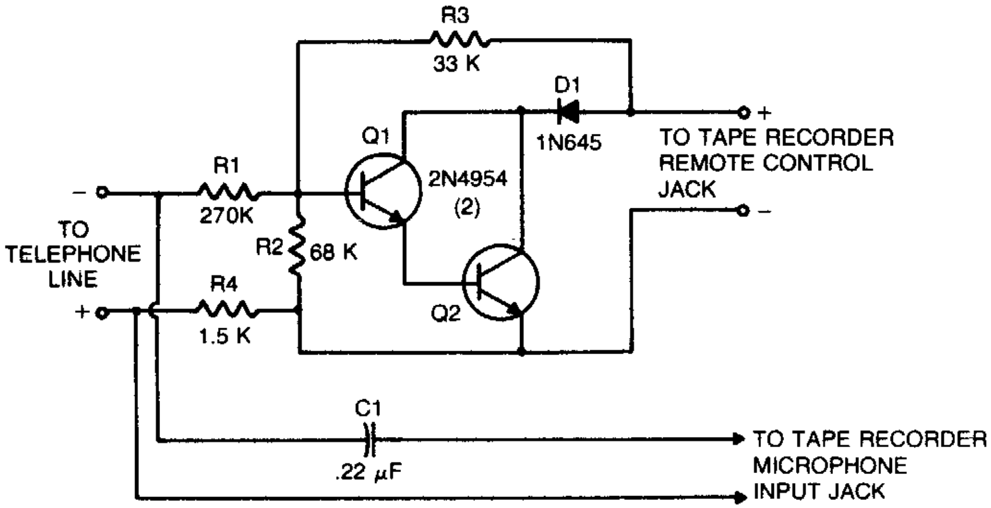


Fig. 85-13

Circuit Notes

This circuit converts a tape recorder into a completely automatic telephone conversation recording instrument that needs no external power source. Voltage at the switch terminals of tape recorder applied to a pair of Darlington-connected transistors, Q1 and Q2, will turn on and start the tape recorder. To turn the transistors off, and thereby stop the machine, apply a negative voltage to the base of Q1 from the phone line. When the telephone receiver is on the hook, there is typically about 50 volts dc across the phone divided across R1, R2, and R4 in such a way that the base of Q1 is sufficiently negative to keep the tape recorder off. When the phone's receiver is picked up, the voltage on the telephone line drops to about 5 volts, which leaves insufficient negative voltage on the base of Q1 to keep it cut off, so the tape recorder starts and begins to record.

TELEPHONE-LINE POWERED REPERTORY DIALER

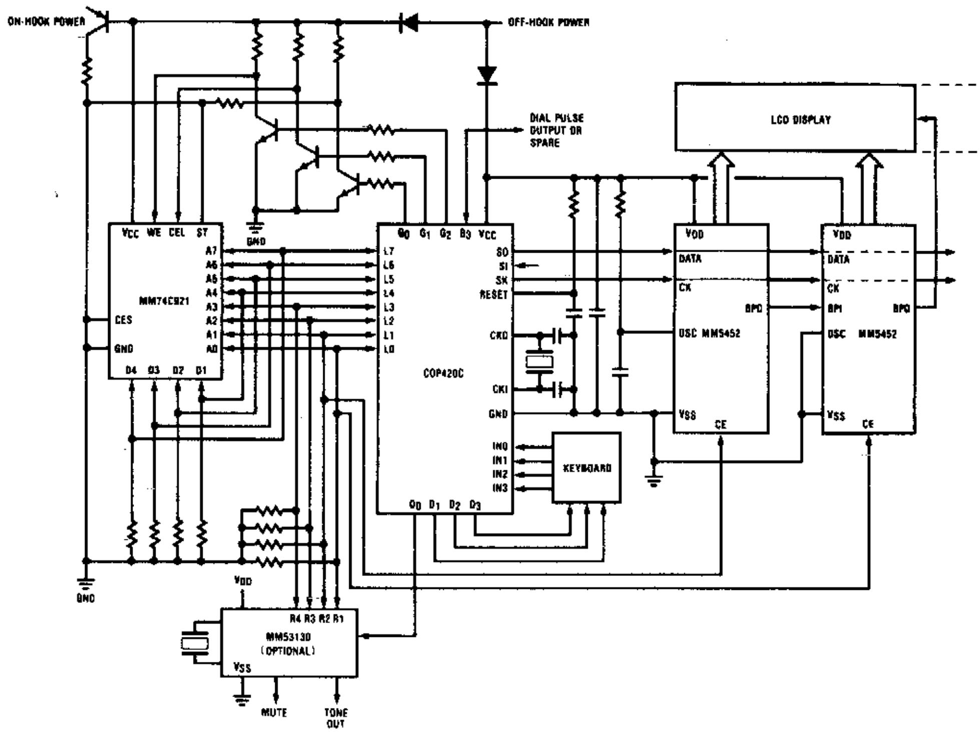


Fig. 85-14

Circuit Notes

Repertory dialer phone has a library of fifteen frequently used numbers, (plus the last number dialed) stored in a standard CMOS RAM. A pushbutton keyboard enables telephone numbers to be keyed in and dialed out directly or a telephone number to be stored in the RAM and dialed automatically.

TELEPHONE OFF-HOOK INDICATOR

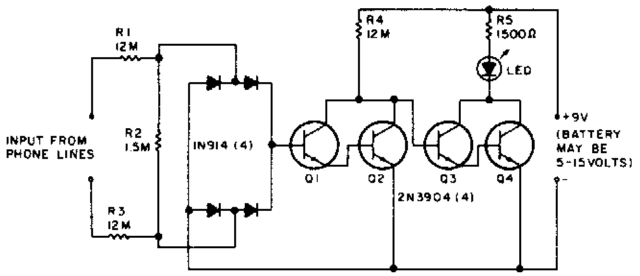


Fig. 85-15

Circuit Notes

The LED flickers when the phone is ringing or being dialed. It glows steadily when the phone is off the hook.

TELEPHONE HANDSET TONE DIAL ENCODER

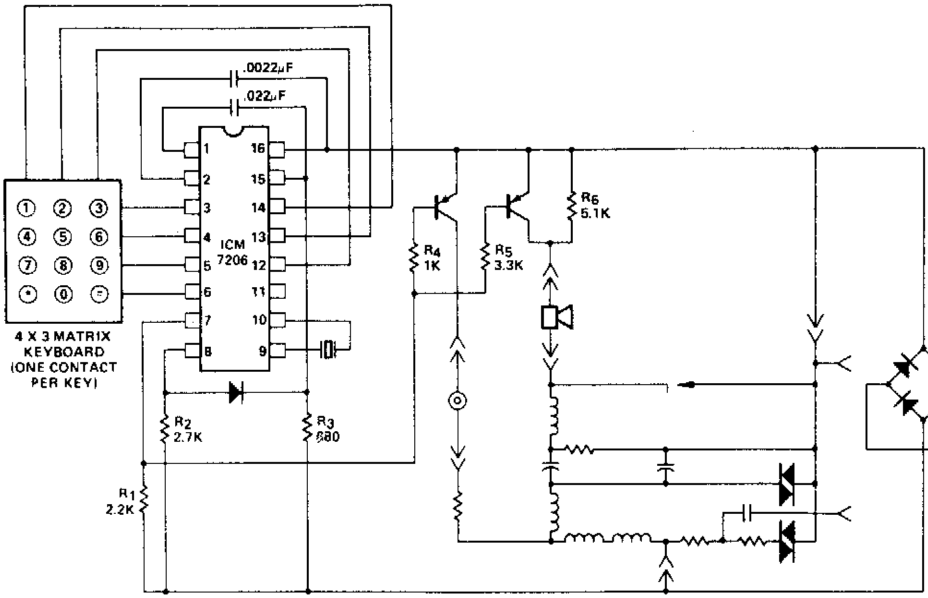


Fig. 85-16

Circuit Notes

This encoder uses a single contact per key keyboard and provides all other switching function electronically. The diode between terminals 8 and 15 prevents the output going more than 1 volt negative with respect to the negative supply V–. The circuit operates over the supply voltage range from 3.5 volts to 15 volts.

LOW LINE LOADING RING DETECTOR

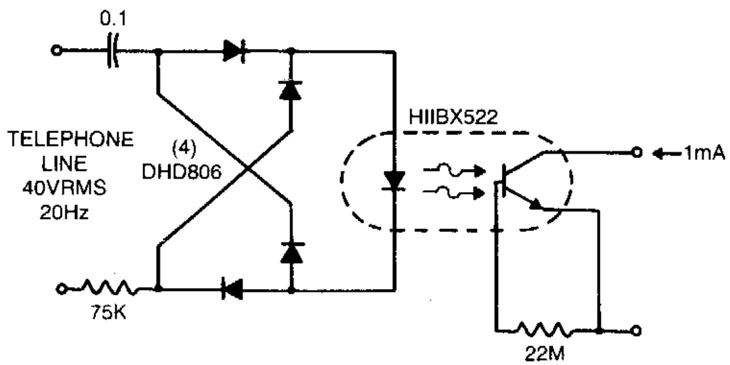
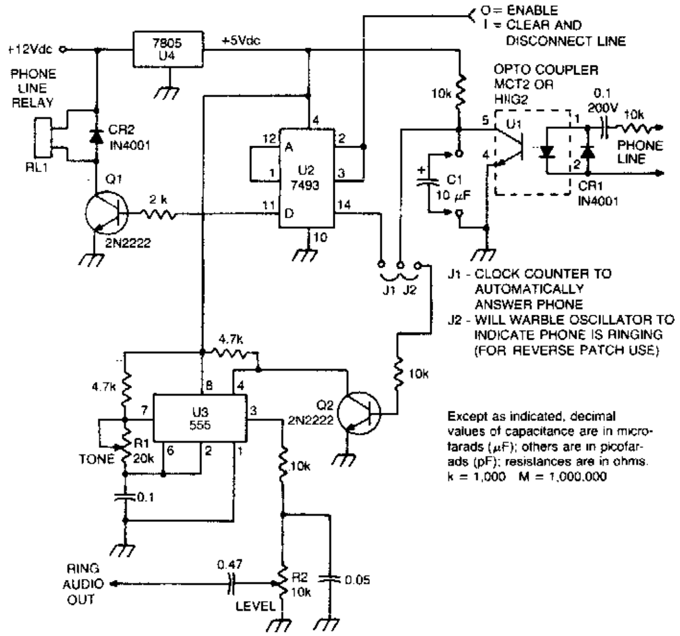


Fig. 85-17

Circuit Notes

Low line current loading is provided by the H11BX522 photodarlington optocoupler, which provides a 1 mA output from a 0.5 mA input.

PHONE AUTO ANSWER AND RING INDICATOR

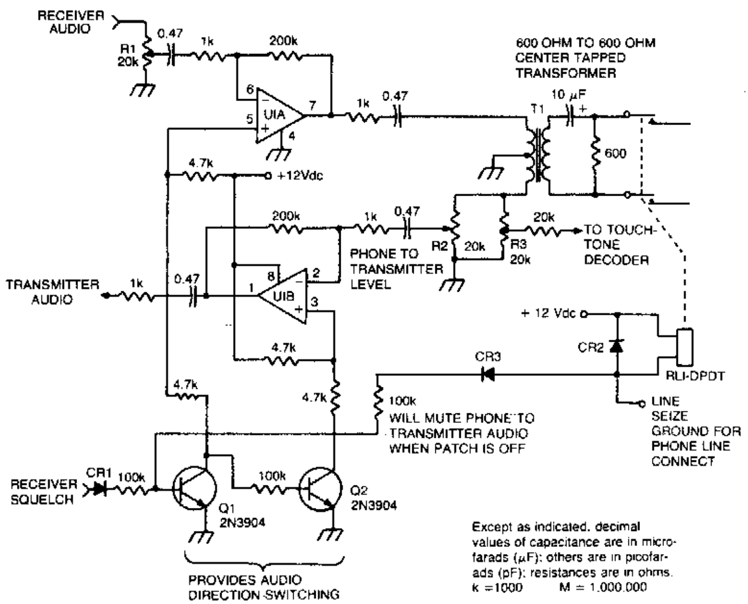


Circuit Notes

Ring detect circuit for automatic phone answering or tone generation for reverse autopatch use.

Fig. 85-18

AUTOPATCH TELEPHONE LINE INTERFACE



Circuit Notes

This circuit provides for the receiver-to-phone line and phone line-to-transmitter link, with both using an op amp for gain.

Fig. 85-19

TELEPHONE RINGER USES PIEZOELECTRIC DEVICE

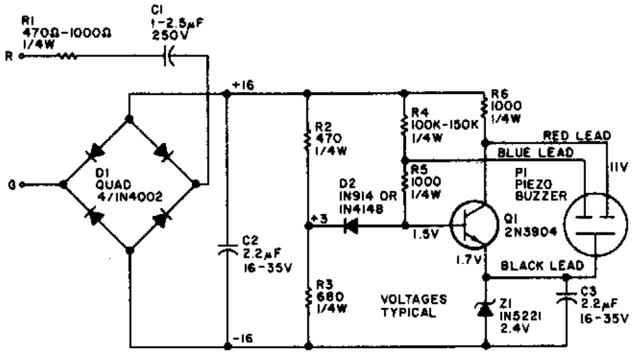


Fig. 85-20

Circait Notes

The electronic bell needs no power supply. Most of the resistors are not critical, although C2, R2, and R3 work best at the values given. Leaving out R1 will make the unit ring louder. The piezo buzzer may vary from store to store. If it has two leads, connect the red lead to the collector and the black lead to the emitter of Q1. If a third (blue) lead is present, connect it to the base of Q1.

ELECTRONIC PHONE BELL

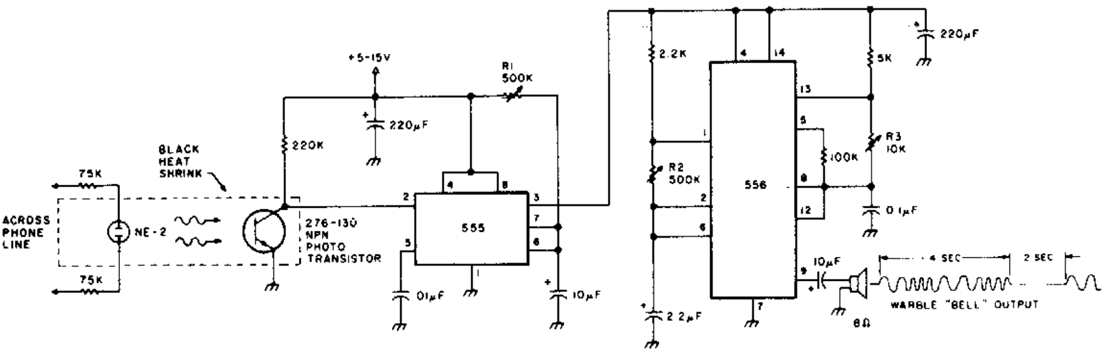


Fig. 85-21

Circuit Notes

The speaker emits a distinctive warble tone when ring pulses are applied to the phone line. Use this circuit as a remote bell or disconnect the phone's ringer for direct use. R1 adjusts the duration of the output; R2 and R3 control the tone's duty cycle and frequency. The transistor is a general-purpose NPN photodevice. The neon bulb and transistor are coupled with the heat-shrink tubing to form an optoisolator.

86

Temperature Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Boiler Control
Heater Control
Two-Wire Remote AC Electronic Thermostat
Three-Wire Electronic Thermostat
Temperature-Sensitive Heater Control
Temperature Controller
Single-Setpoint Temperature Controller
Temperature Controller
Temperature Control
Temperature Controller
Temperature Controller
Portable Calibrator

BOILER CONTROL

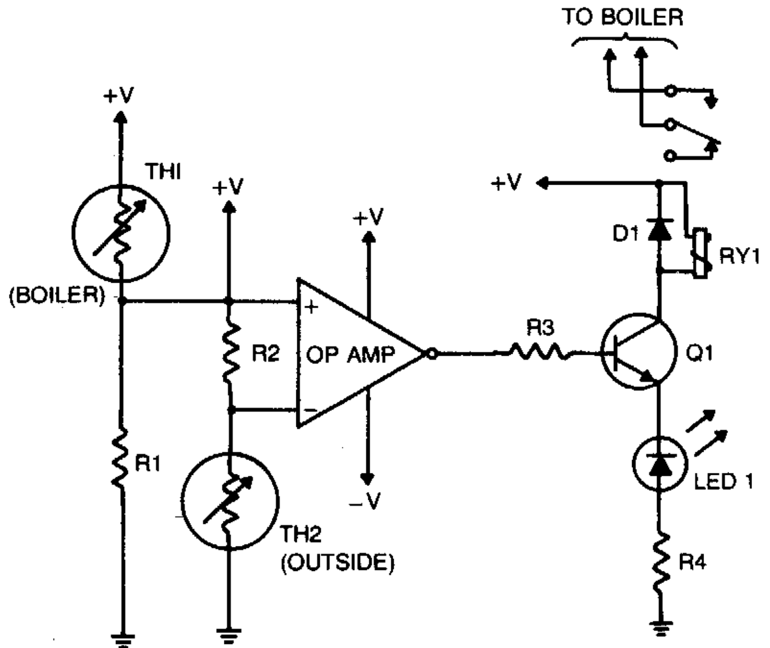
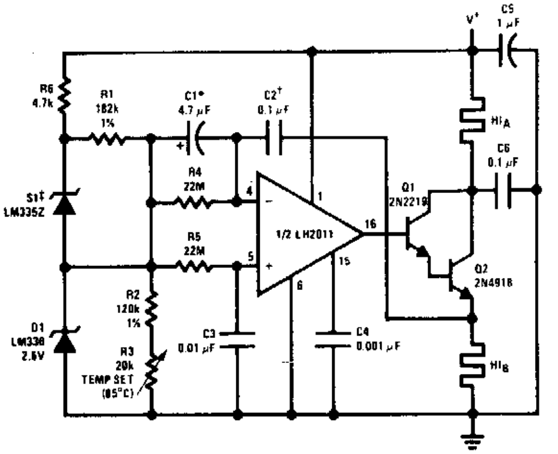


Fig. 86-1

Circuit Notes

The purpose of this circuit is to control the water temperature in a hot-water heating system. What it does is to lower the boiler temperature as the outside air temperature increases. The op amp is used as a comparator. Thermistor TH2 and R2 form a voltage divider that supplies a reference voltage to the op-amp's inverting input. Thermistor TH2 is placed outdoors, and the values of TH2 and R2 should be chosen so that when the outside temperature is 25 °F, the resistance of the thermistor and resistor are equal. Resistor R1 and thermistor TH1 make up a voltage divider that supplies a voltage to the op amp's noninverting input. Thermistor TH1 is placed inside the boiler and the values of TH1 and R1 should be chosen so that when the boiler's temperature is 160 °F, their resistances are equal. The output of the op amp controls Q1, which is configured as a transistor switch. When the logic output of the op amp is high, Q1 is turned on, energizing relay RY1. The relay's contacts should be wired so that the boiler's heat supply is turned off (relay energized).

HEATER CONTROL



* solid tantalum

† mylar

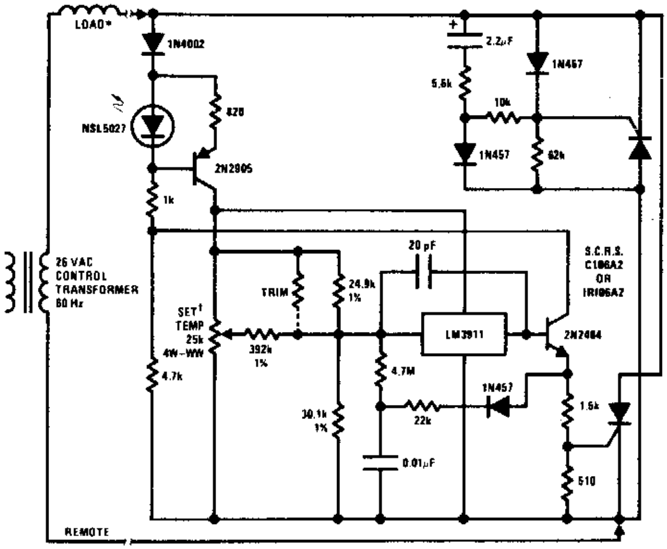
‡ close thermal coupling between sensor and oven shell is recommended.

Circuit Notes

This proportional control crystal oven heater uses lead/lag compensation for fast setting. The time constant is changed with R4 and compensating resistor R5. If Q2 is inside the oven, a regulated supply is recommended for 0.1 °C. control.

Fig. 86-2

TWO-WIRE REMOTE AC ELECTRONIC THERMOSTAT (GAS OR OIL FURNACE CONTROL)



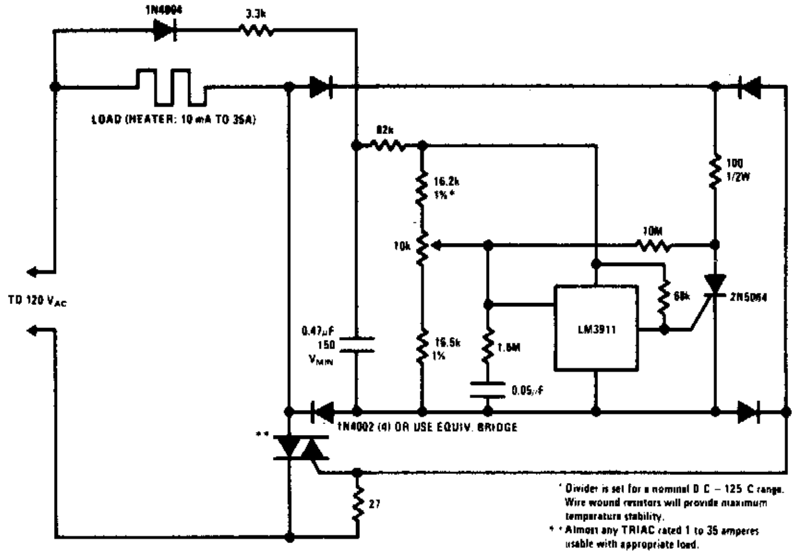
* Solenoid or 6–15W heater

† Pot will provide about a 50°F to 90°F setting range. The trim resistor (100k) is selected to bring 70°F near the middle of the pot rotation.

SCR heating, by proper positioning, can preheat the sensor giving control anticipation as is presently used in many home thermostats.

Fig. 86-3

THREE-WIRE ELECTRONIC THERMOSTAT



* Divider is set for a nominal 0 C – 125 C range. Wire wound resistors will provide maximum temperature stability.

** Almost any TRIAC rated 1 to 35 amperes usable with appropriate load.

Fig. 86-4

TEMPERATURE-SENSITIVE HEATER CONTROL

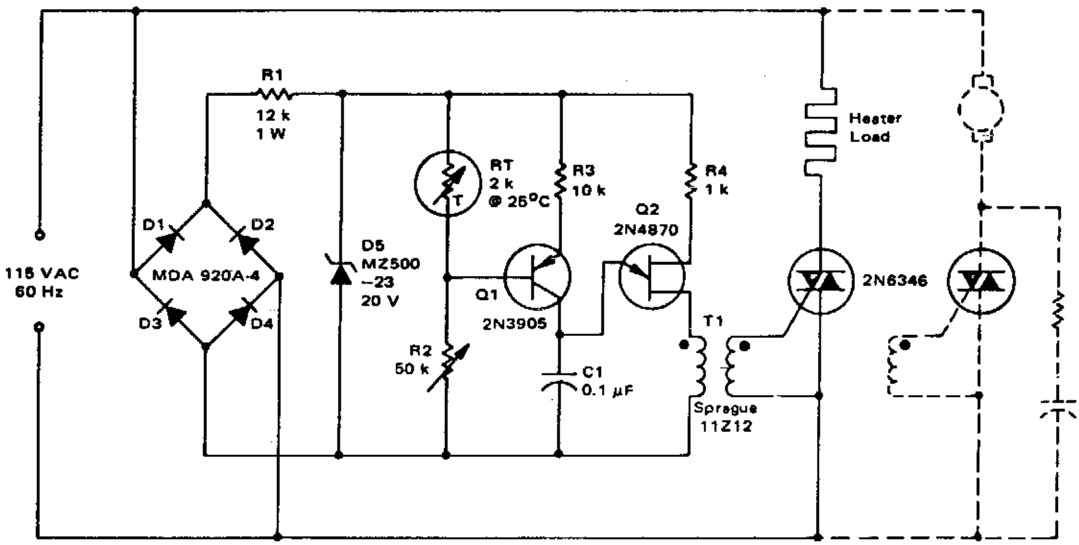


Fig. 86-5

TEMPERATURE CONTROLLER

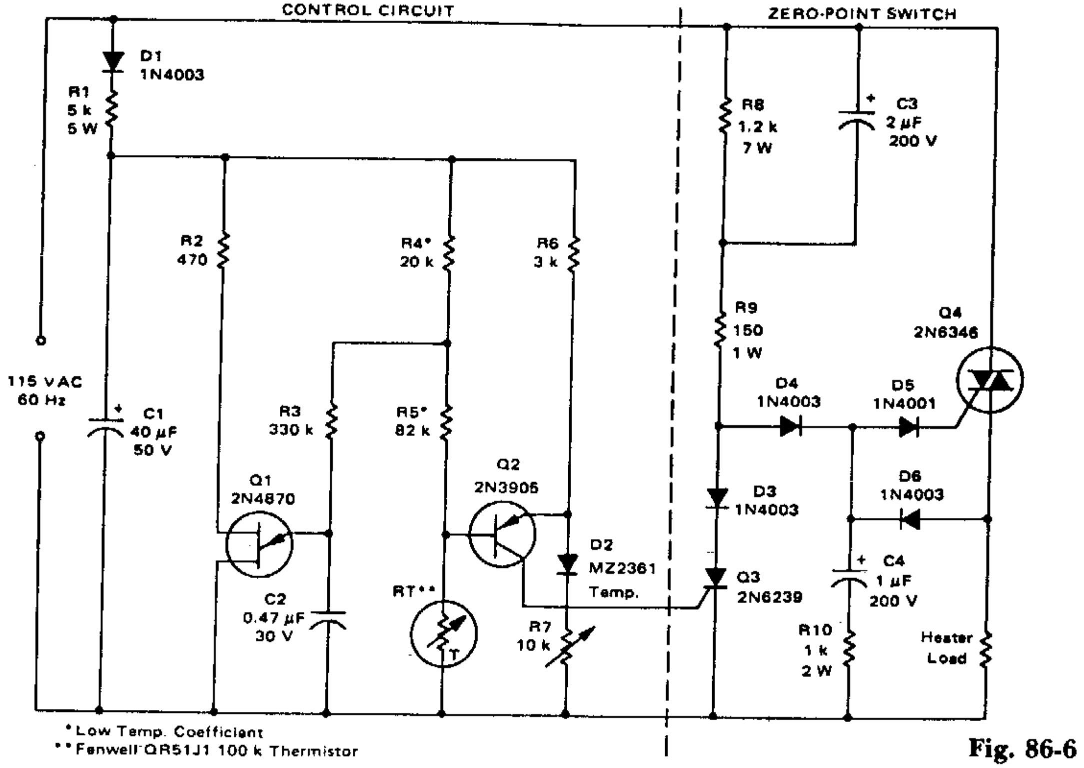


Fig. 86-6

SINGLE-SETPOINT TEMPERATRE CONTROLLER

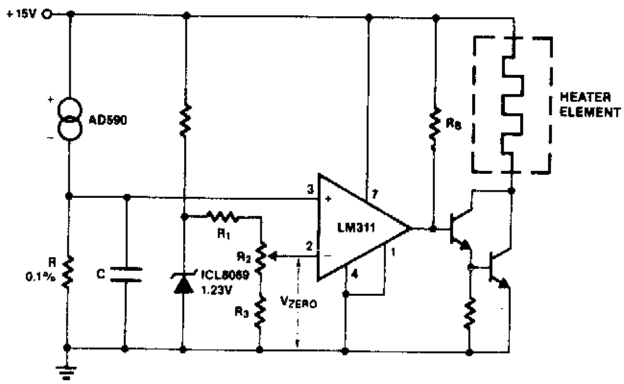


Fig. 86-7

Circuit Notes

The AD590 produces a temperature-dependent voltage across R (C is for filtering noise). Setting R2 produces a scale-zero voltage. For the Celsius scale, make R = 1 K and V_{zero} = 0.273 volts. For Fahrenheit, R = 1.8 K and V_{zero} = 0.460 volts.

TEMPERATURE CONTROLLER

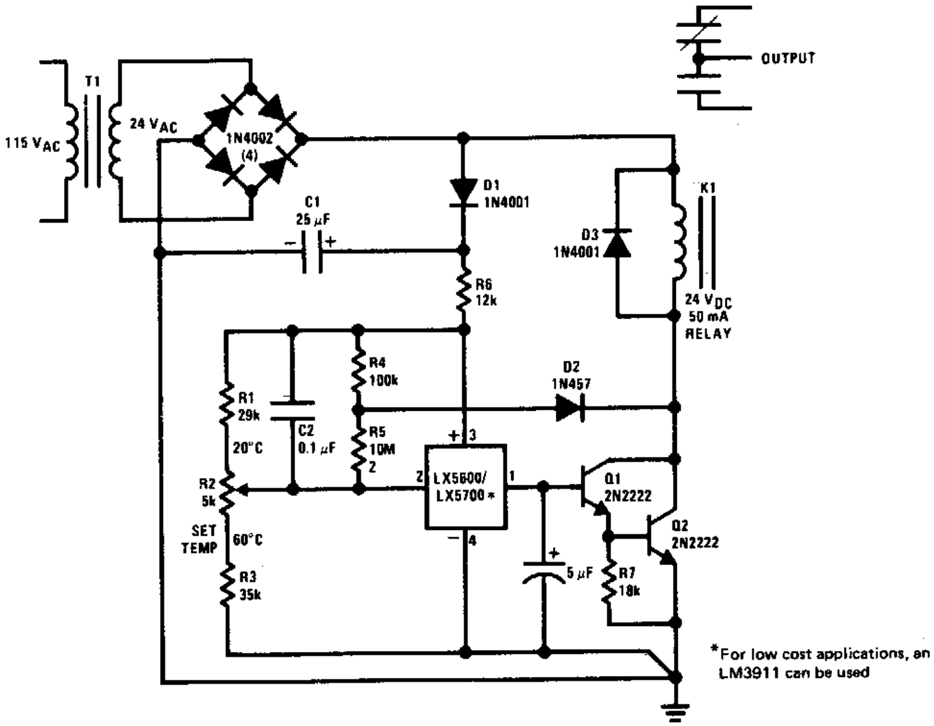


Fig. 86-8

Circuit Notes

The sensor is a standard TO-5 or TO-46 package. For surface or air temperature sensing. Small clip-on heat sinks can be used. A simple probe can be made using heat-shrink tubing and RTV silicon rubber. Three-leads-plus-shield cable is a good choice for wire with the shield connected to pin 4. The controller can be used for baths, ovens, oven-temperature protection, or even home thermostats. Long-term stability and repeatability is better than 0.5 °C.

TEMPERATURE CONTROL

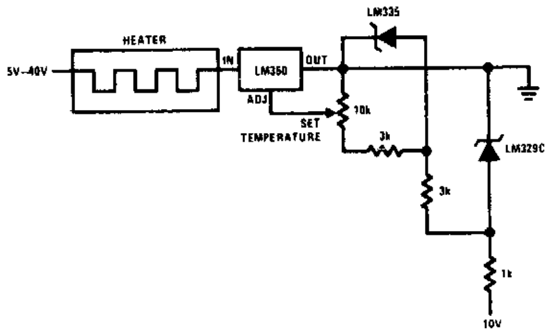


Fig. 86-9

TEMPEATURE CONTROLLER

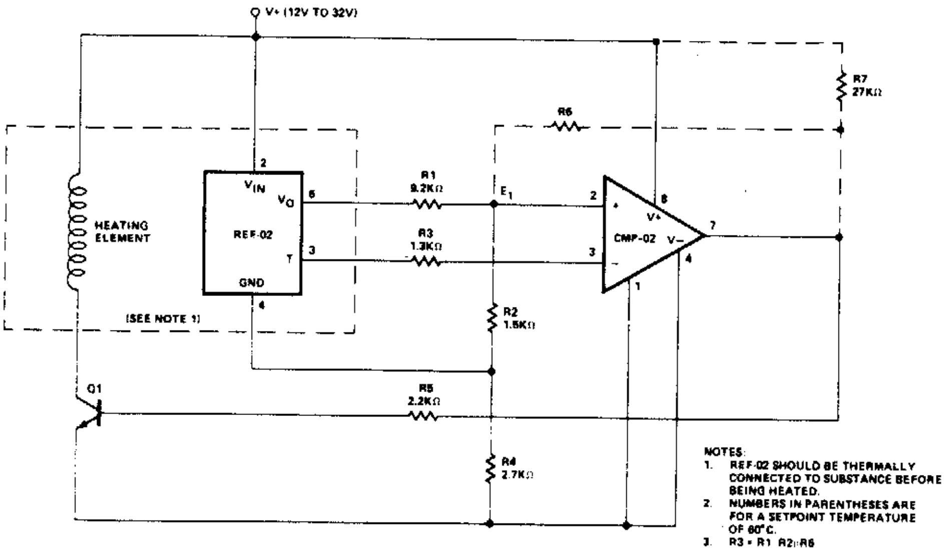


Fig. 86-10

Circuit Notes

Temperature control is achieved using the REF-02 +5 V Reference/Thermometer and a CMP-02 Precision Low Input Current Comparator. The CMP-02 turns on a heating element driver (Q1) whenever the present temperature drops below a setpoint temperature determined by the ratio of R1 to R2. The circuit also provides adjustable hysteresis and single supply operation.

TEMPERATURE CONTROLLER

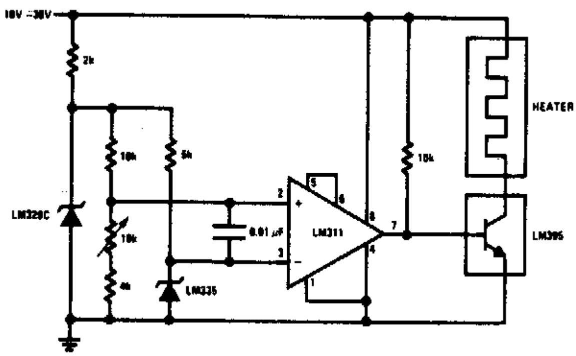
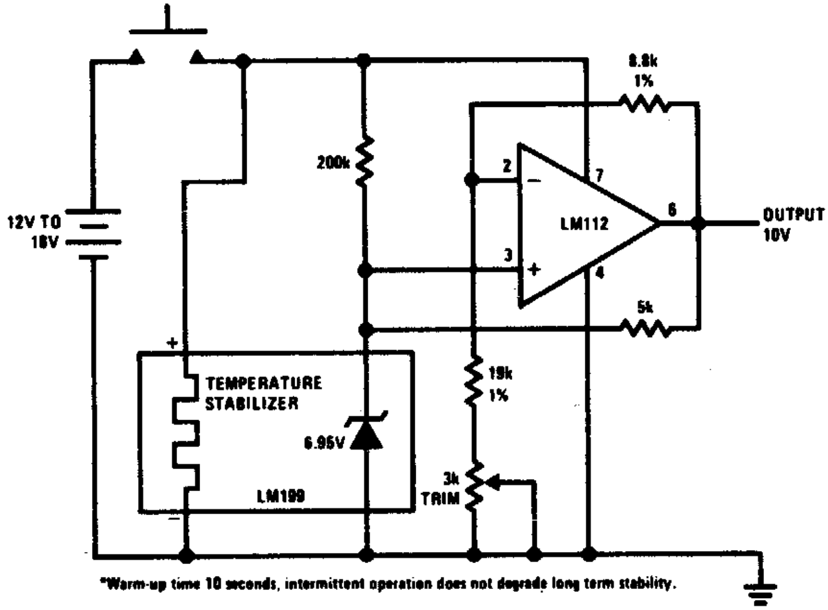


Fig. 86-11

PORTABLE CALIBRATOR



*Warm-up time 10 seconds, intermittent operation does not degrade long term stability.

Fig. 86-12

87

Temperature Sensors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LINEAR TEMPERATURE-TO-FREQUENCY TRANSCONDUCER

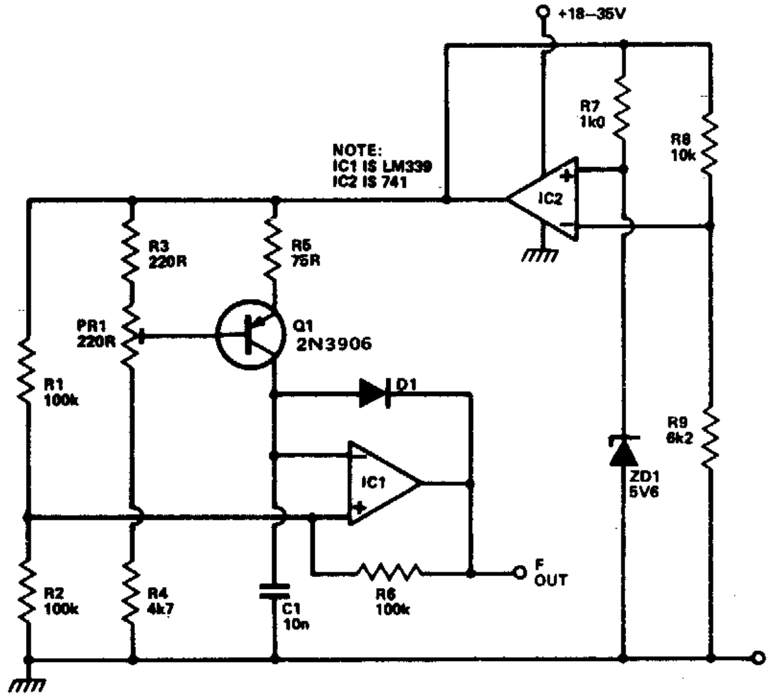


Fig. 87-1

Circuit Notes

This circuit provides a linear increase of frequency of 10 Hz/°C over 0-100 °C and can thus be used with logic systems, including microprocessors. Temperature probes Q1 V_{be} changes 2.2 mV/°C. This transistor is incorporated in a constant current source circuit. Thus, a current proportional to temperature will be available to charge C1. The circuit is powered via the temperature stable reference voltage supplied by the 741. Comparator IC1 is used as a Schmitt trigger whose output is used to discharge C1 via D1. To calibrate the circuit Q1 is immersed in boiling distilled water and PR1 adjusted to give 1 kHz output. The prototype was found to be accurate to within 0.2 °C.

TEMPERATURE METER

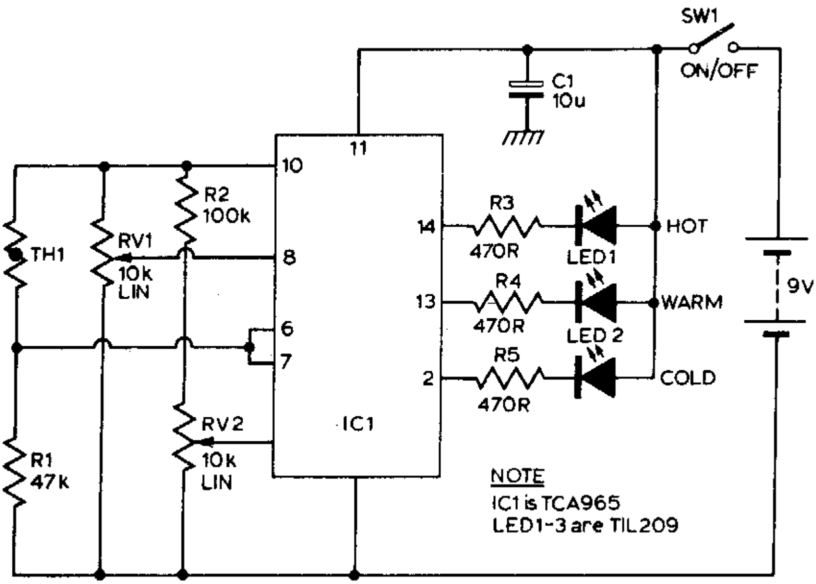


Fig. 87-2

Circuit Notes

TCA965 window discriminator IC allows the potentiometers RV1 and RV2 to set up a window height and window width respectively. R1 and thermistor TH1 for a potential divider connected across the supply lines. R1 is chosen such that at ambient temperature the voltage at the junction of these two components will be approximately half supply. As the temperature of the sensor changes, the voltage will change. RV1 will set the point which corresponds to the center voltage of a window the width of which is set by RV2. The switching points of the IC feature a Schmitt characteristic with low hysteresis. The outputs of IC1 indicate whether the input voltage is within the window or outside by virtue of being either too high or too low. The outputs of IC1 drive the LEDs via a current limiting resistor.

FOUR-CHANNEL TEMPERATURE SENSOR (0-50 °C)

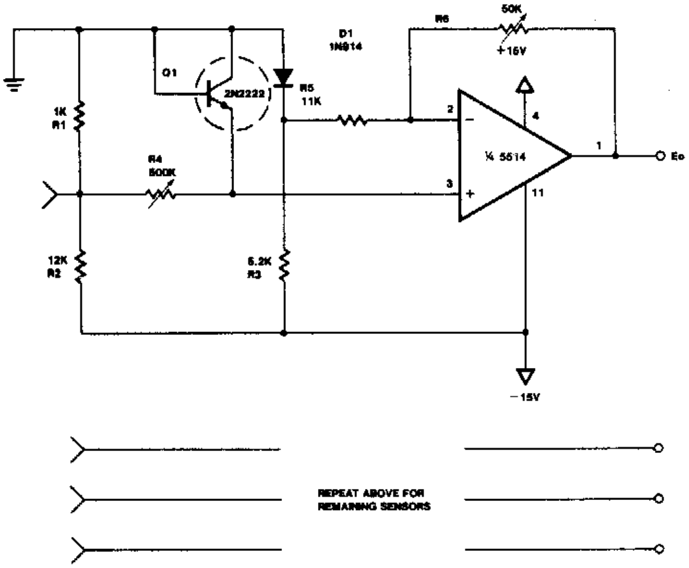
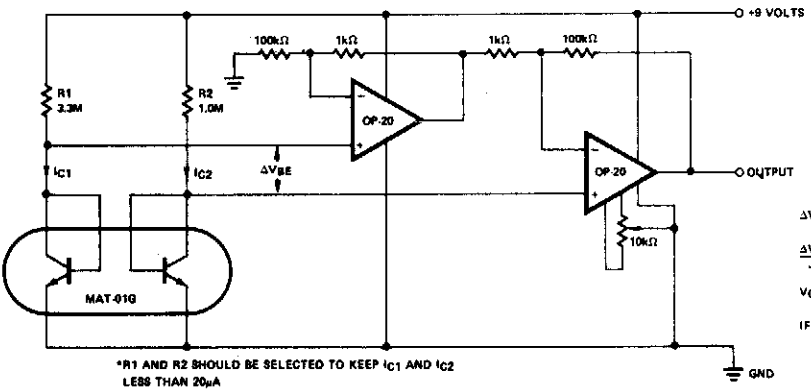


Fig. 87-3

TEMPERATURE SENSOR



DESIGN EQUATIONS

$\Delta V_{BE} = \frac{kT}{q} \text{LN} \left(\frac{I_{C2}}{I_{C1}}\right)$

$\frac{\Delta V_{BE}}{\Delta T} = 86.8 \text{ LN} \left(\frac{I_{C2}}{I_{C1}}\right) (\mu V/°K)$

$V_{OUT} = 101(\Delta V_{BE})$

IF $\frac{R1}{R2} = 3.2$ THEN $TCV_{OUT} = 10mV/°C$

Fig. 87-4

INTEGRATED CIRCUIT TEMPERATURE SENSOR

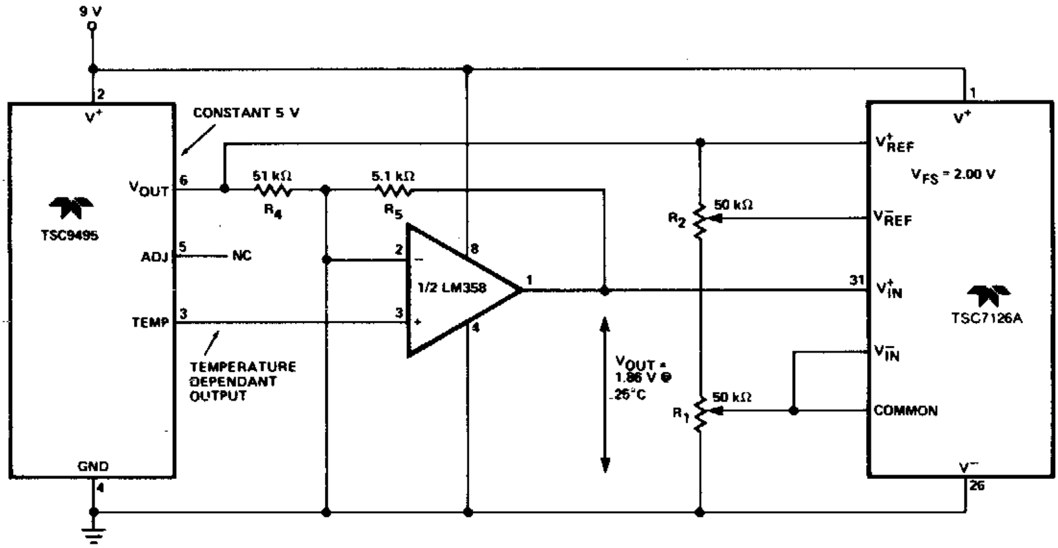
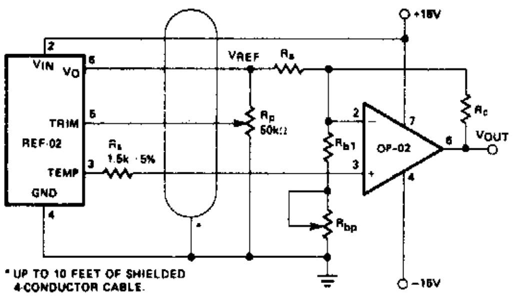


Fig. 87-5

PRECISION TEMPERATURE TRANSDUCER WITH REMOTE SENSOR



* UP TO 10 FEET OF SHIELDED 4-CONDUCTOR CABLE.

FOR THEORY OF OPERATION AND CALIBRATION PROCEDURE CONSULT APPLICATION NOTE 18, "THERMOMETER APPLICATIONS OF THE REF-02".

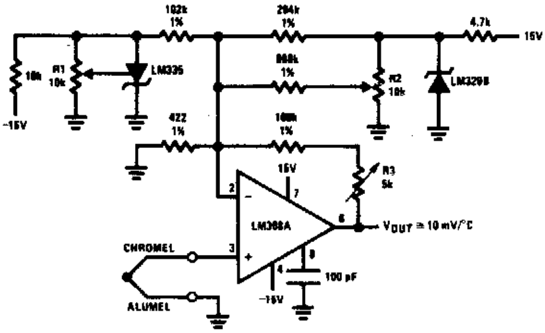
RESISTOR VALUES

TCV_{OUT} SLOPE (S)	10mV/°C	100mV/°C	10mV/°F
TEMPERATURE RANGE	−55°C to +125°C	−55°C to +125°C	−67°F to +257°C
OUTPUT VOLTAGE RANGE	−0.55V to +1.25V	−5.5V to +12.5V	−0.67V to +2.57V
ZERO SCALE	0V@0°C	0V@0°C	0V@0°F
R_a (±1% resistor)	9.09kΩ	15kΩ	7.5kΩ
R_{b1} (±% resistor)	1.5kΩ	1.82kΩ	1.21kΩ
R_{bp} (Potentiometer)	200Ω	500Ω	200Ω
R_c (±1% resistor)	5.11kΩ	84.5kΩ	8.25kΩ

* For 125°C operation, the op amp output must be able to swing to +12.5V, increase V_{IN} to +18V from +15V if this is a problem.

Fig. 87-6

CENTIGRADE CALIBRATED THERMOCOUPLE THERMOMETER



Terminate thermocouple reference junction in close proximity to LM335.

Adjustments:

1. Apply signal in place of thermocouple and adjust R3 for a gain of 245.7.
2. Short non-inverting input of LM308A and output of LM329B to ground.
3. Adjust R1 so that V_{OUT} = 2.982V @ 25°C.
4. Remove short across LM329B and adjust R2 so that V_{OUT} = 246 mV @ 25°C.
5. Remove short across thermocouple.

Fig. 87-7

μP CONTROLLED DIGITAL THERMOMETER

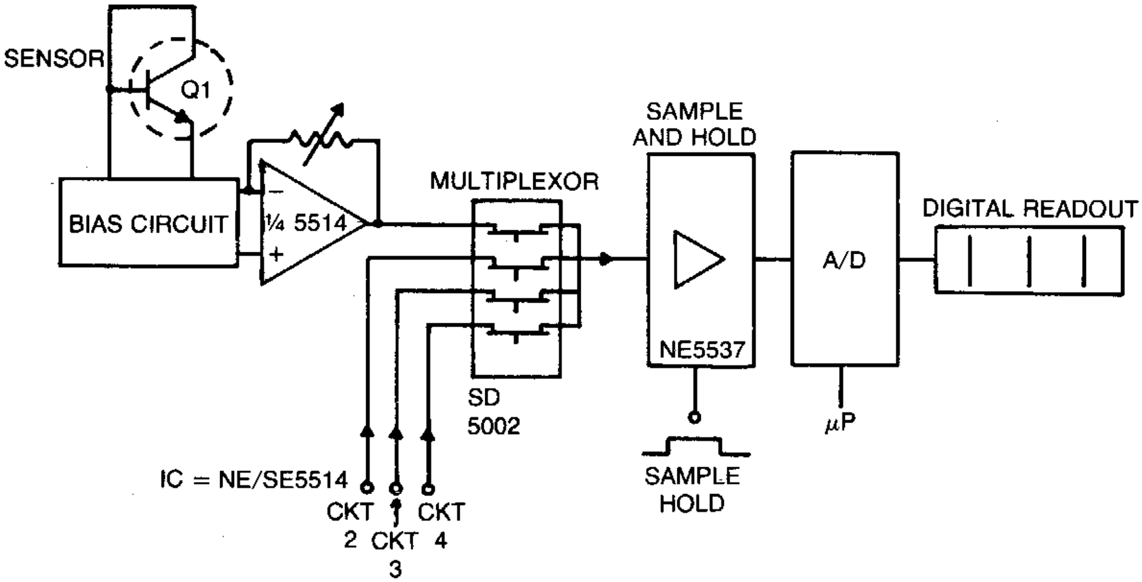


Fig. 87-8

ISOLATED TEMPERATURE SENSOR

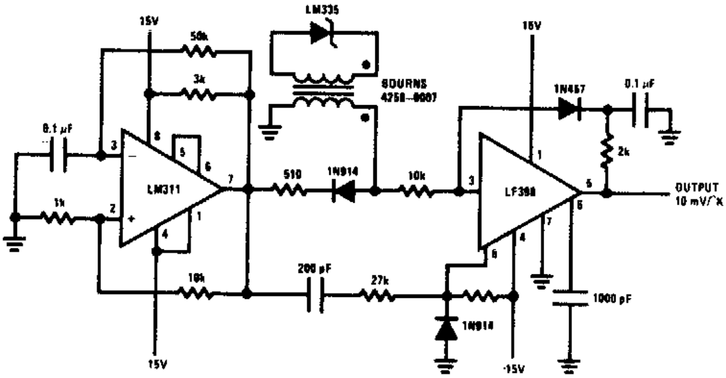


Fig. 87-9

DIGITAL THERMOMETER

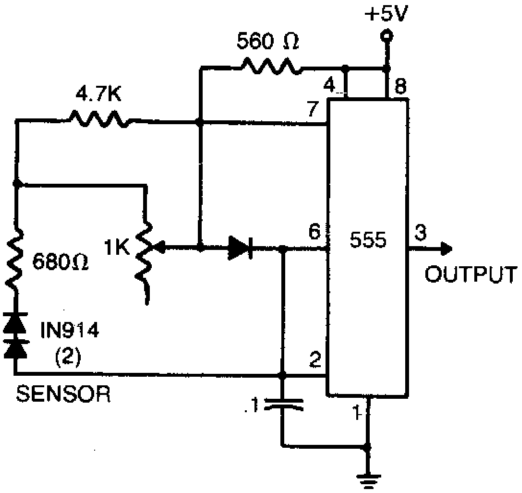


Fig. 87-10

Circuit Notes

The sensor consists of two series-connected 1N914s, part of the circuit of a 555 multivibrator. Wired as shown, the output pulse rate is proportional to the temperature of the diodes. This output is fed to a simple frequency-counting circuit.

VARIABLE OFFSET THERMOMETER

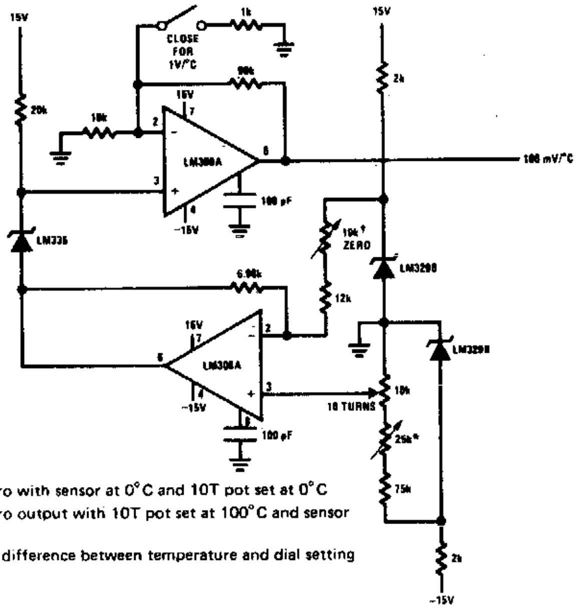


Fig. 87-11

† Adjust for zero with sensor at 0°C and 10T pot set at 0°C

* Adjust for zero output with 10T pot set at 100°C and sensor at 100°C

‡ Output reads difference between temperature and dial setting of 10T pot

DIFFERENTIAL THERMOMETER

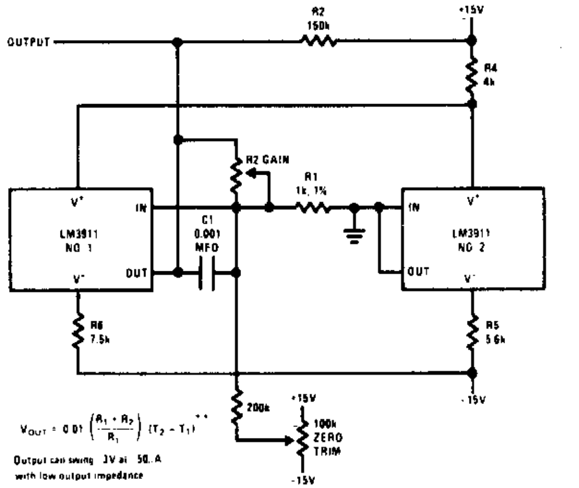


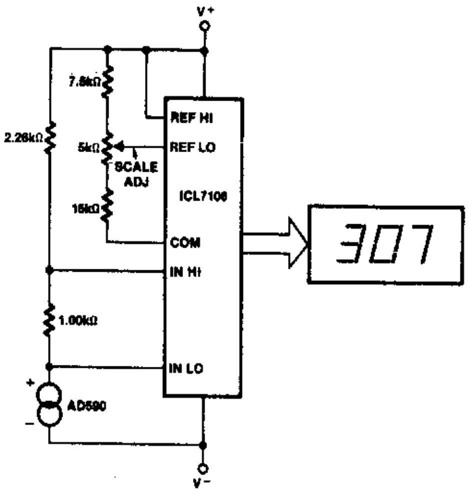
Fig. 87-12

$$V_{OUT} = 0.01 \left(\frac{R_1 + R_2}{R_1}\right) (T_2 - T_1)^{**}$$

Output can swing 3V at 50 μA with low output impedance

** The 0.01 in the above equation is in units of V/°K or V/°C, and is a result of the basic 0.01 V/°K sensitivity of the transducer

BASIC DIGITAL THERMOMETER, KELVIN SCALE



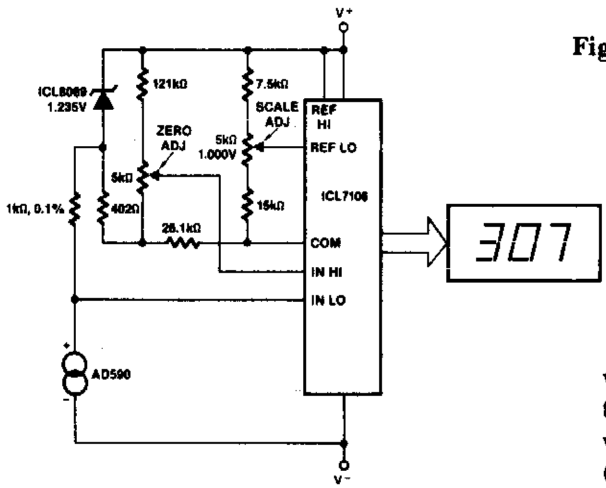
Circuit Notes

The Kelvin scale version reads from 0 to 1999 °K theoretically, and from 223 °K to 473 °K actually. The 2.26 K resistor brings the input within the ICL7106 V_{CM} range: two general-purpose silicon diodes or an LED may be subsituted.

Fig. 87-13

BASIC DIGITAL THERMOMETER, KELVIN SCALE WITH ZERO ADJUST

Fig. 87-14



Circuit Notes

This circuit allows zero adjustment as well as slope adjustment. The ICL8069 brings the input within the common-mode range, while the 5 K pots trim any offset at 218 °K (−55 °C), and set scale factor.

THERMOCOUPLE AMPLIFIER

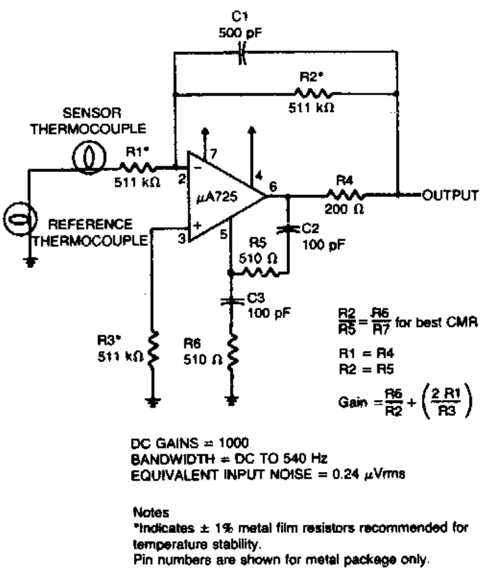


Fig. 87-15

REMOTE TEMPERATURE SENSING

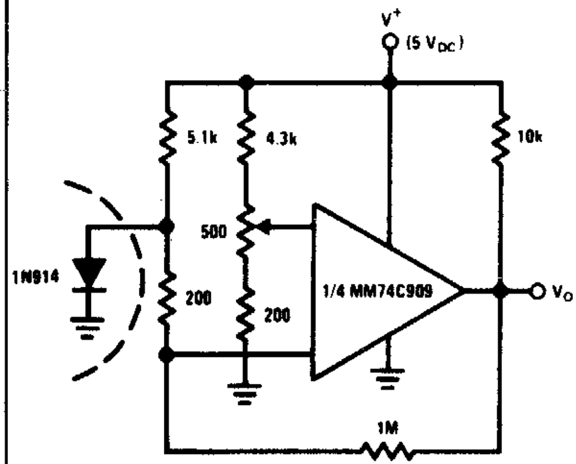


Fig. 87-17

OPTICAL PYROMETER

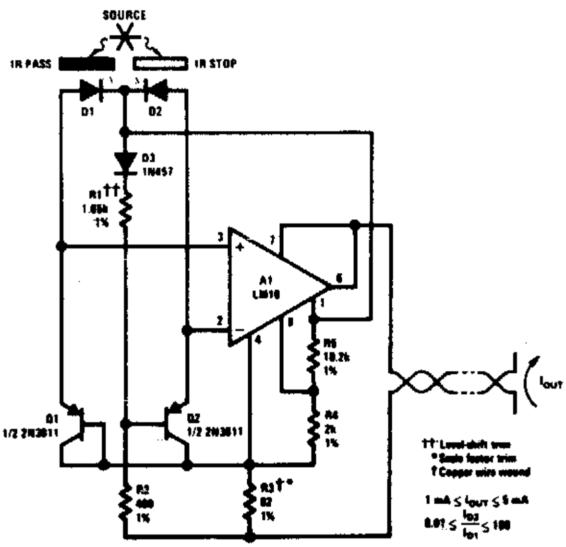


Fig. 87-16

SIMPLE DIFFERENTIAL TEMPERATURE SENSOR

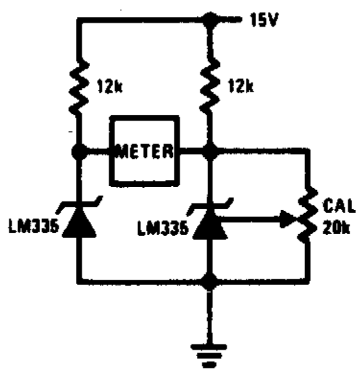


Fig. 87-18

DIFFERENTIAL TEMPERATURE SENSOR

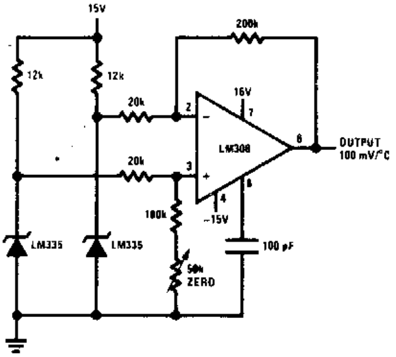


Fig. 87-19

KELVIN THERMOMETER WITH GROUND REFERRED OUTPUT

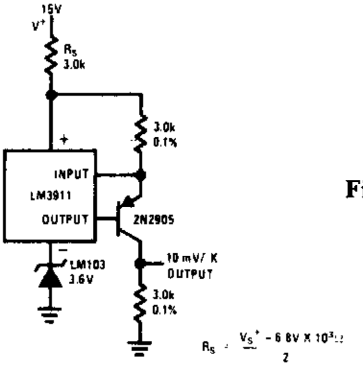
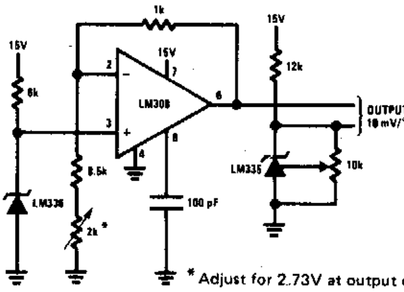


Fig. 87-22

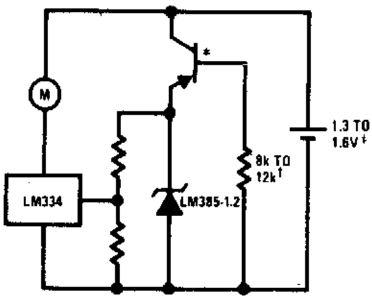
CENTIGRADE THERMOMETER



* Adjust for 2.73V at output of LM308

Fig. 87-20

LOWER POWER THERMOMETER



* 2N3638 or 2N2907 select for inverse $H_{FE} \cong 5$

† Select for operation at 1.3V

‡ $I_Q \cong 600\ \mu A$ to $900\ \mu A$

Fig. 87-23

METER THERMOMETER WITH TRIMMED OUTPUT

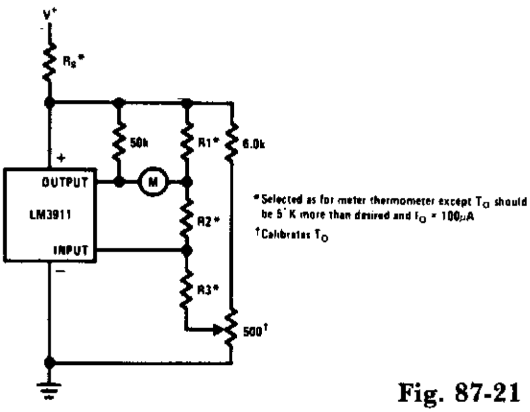


Fig. 87-21

0 °F-50 °F THERMOMETER

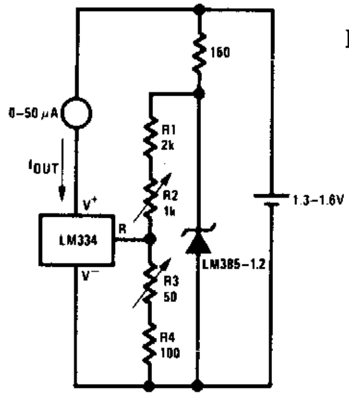


Fig. 87-24

Calibration

1. Short LM385-1.2, adjust R3 for I_{OUT} = temp at 1.8 $\mu A/°K$
2. Remove short, adjust R2 for correct reading in °F

TEMPERATURE-TO-FREQUENCY CONVERTER

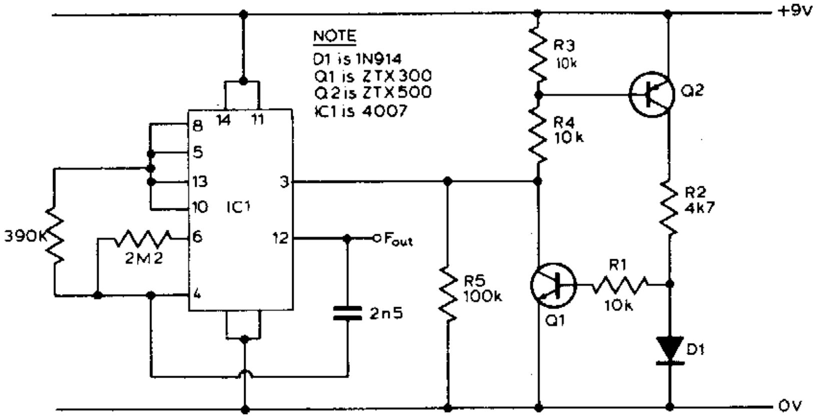
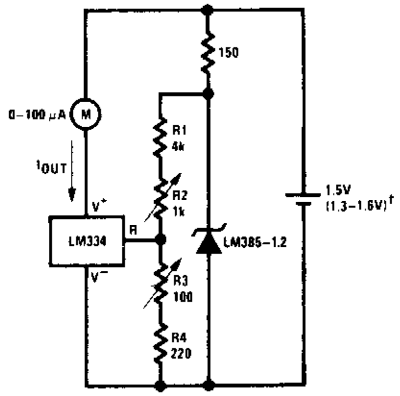


Fig. 87-25

Circuit Notes

The circuit exploits the fact that when fed from a constant current source, the forward voltage of a silicon diode varies with temperature in a reasonably linear way. Diode D1 and resistor R2 form a potential divider fed from the constant current source. As the temperature rises, the forward voltage of D1 falls tending to turn Q1 off. The output voltage from Q1 will thus rise, and this is used as the control voltage for the CMOS VCO. With the values shown, the device gave an increase of just under 3 Hz/°C (between 0 °C and 60 °C) giving a frequency of 470 Hz at 0 °C.

0 °C-100 °C THERMOMETER



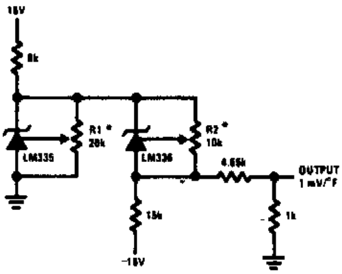
Calibration

1. Short LM385-1.2, adjust R3 for I_{OUT} = temp at 1 μA/°K
2. Remove short, adjust R2 for correct reading in centigrade

†I_Q at 1.3V ≅ 500 μA
I_Q at 1.6V ≅ 2.4 mA

Fig. 87-26

GROUND REFERRED FAHRENHEIT THERMOMETER



*Adjust R2 for 2.554V across LM336.
Adjust R1 for correct output.

Fig. 87-27

GROUND REFERRED CENTIGRADE THERMOMETER

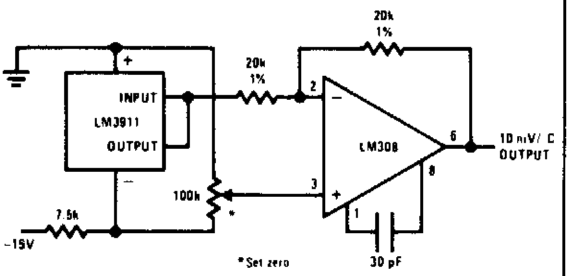


Fig. 87-28

TEMPERATURE SENSOR

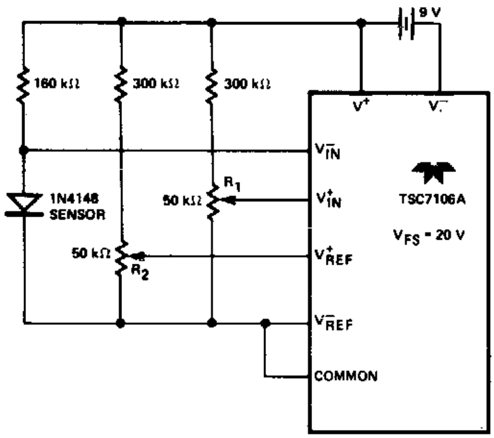


Fig. 87-30

GROUND REFERRED CENTIGRADE THERMOMETER

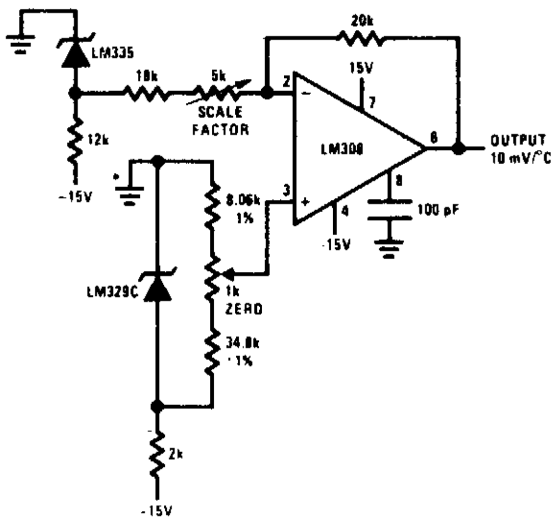


Fig. 87-29

POSITIVE TEMPERATURE SENSOR COEFFICIENT RESISTOR

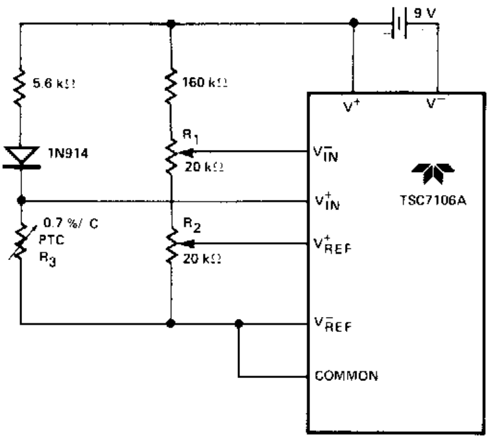
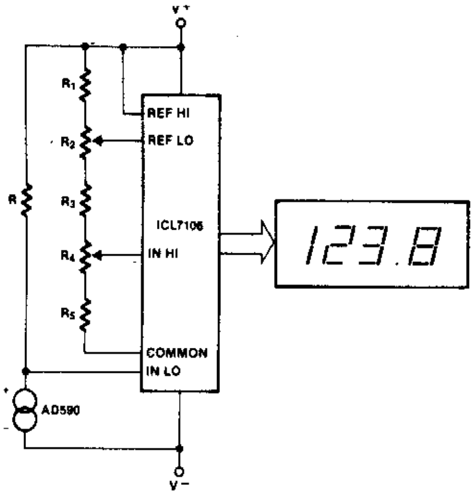


Fig. 87-31

BASIC DIGITAL THERMOMETER (CELSIUS AND FAHRENHEIT SCALES)



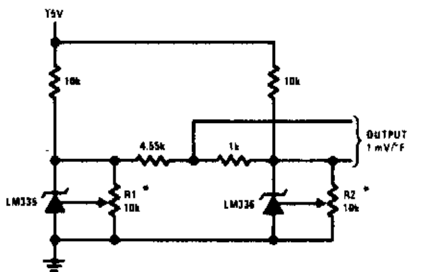
	R	R_1	R_2	R_3	R_4	R_5
°F	9.00	4.02	2.0	12.4	10.0	0
°C	5.00	4.02	2.0	5.11	5.0	11.8

Fig. 87-32

Circuit Notes

Maximum reading on the Celsius range is 199.9 °C, limited by the (short-term) maximum allowable sensor temperature. Maximum reading on the Fahrenheit range is 199.9 °F (93.3 °C), limited by the number of display digits. V_{REF} for both scales is 500 mV.

FAHRENHEIT THERMOMETER



*To calibrate adjust R2 for 2.554V across LM336.
Adjust R1 for correct output.

Fig. 87-33

88

Timers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

THUMBWHEEL PROGRAMMABLE INTERVAL TIMER

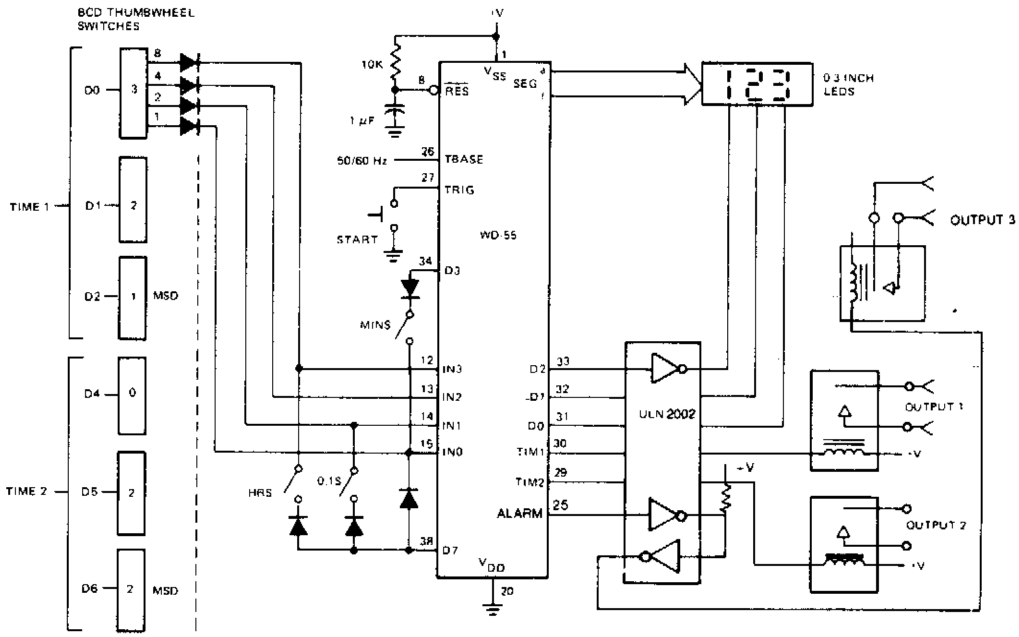
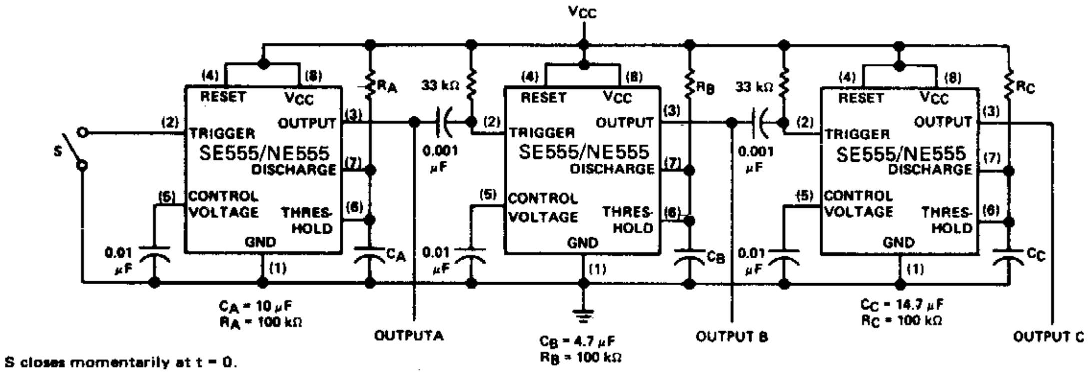


Fig. 88-1

Circuit Notes

Switch programmable on/off or interval timer, has three relay-switched outputs. Output one is active for the duration of time 1, output two is active for the duration of time 2, and output three is active for the duration of both one and two. Timing data is input through 6 BCD-encoded thumbwheel switches. Three SPST switches inform the WD-55 to interpret this data as NNN seconds. NNN seconds, NNN minutes, or NNN hours. The LED display will show the time remaining and the countdown when operating. Since the data is input through switches, the display may be deleted. Also, since the timing information is read from switches, the data is nonvolatile and no battery backup is required.

SEQUENTIAL TIMER



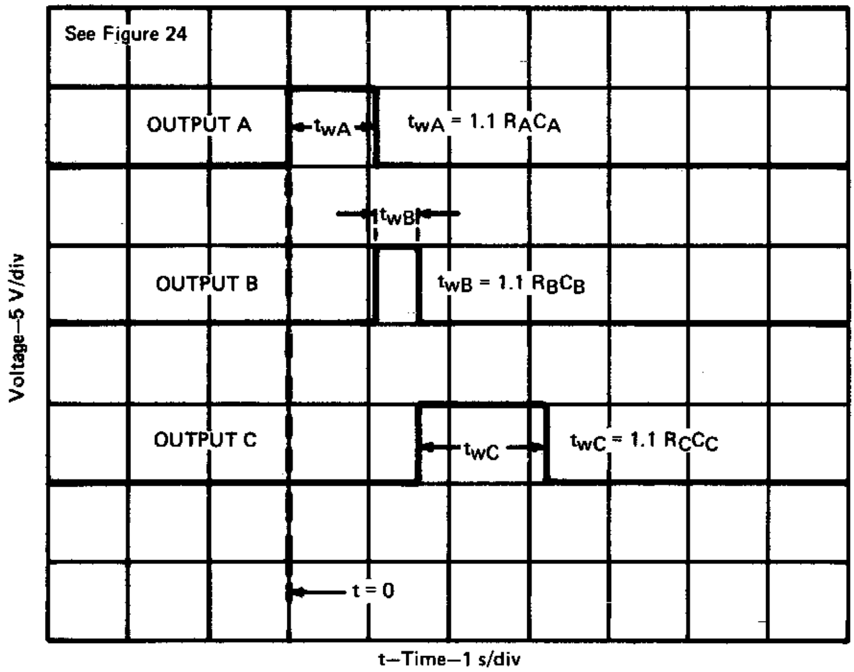


Fig. 88-2

Circuit Notes

Many applications, such as computers, require signals for initializing conditions during start-up. Other applications such as test equipment require activation of test signals in sequence. SE555/NE555 circuits may be connected to provide such sequential control. The timers may be used in various combinations of astable or monostable circuit connections, with or without modulation, for extremely flexible waveform control.

SEQUENTIAL TIMER

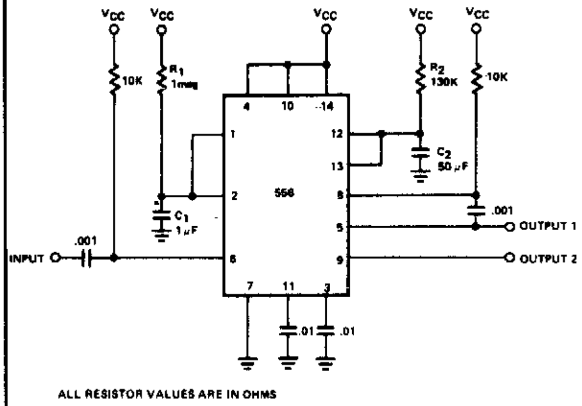


Fig. 88-3

Circuit Notes

By utilizing both halves of a dual timer it is possible to obtain sequential timing. By connecting the output of the first half to the input of the second half via a .001 µF coupling capacitor sequential timing may be obtained. Delay t_1 is determined by the first half and t_2 by the second half delay. The first half of the timer is started by momentarily connecting pin 6 to ground. When it is turned out (determined by 1.1R1C1), the second half begins. Its duration is determined by 1.1R2C2.

SEQUENTIAL UJT TIMER

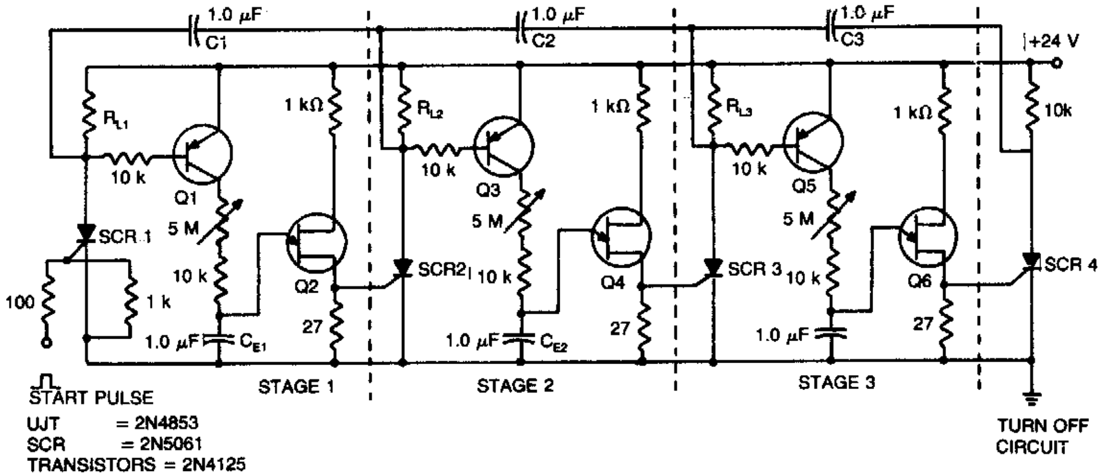


Fig. 88-4

TIME-DELAYED RELAY (FOR PATIO-LIGHT, GARAGE LIGHT, ENLARGER PHOTOTIMER, ETC.)

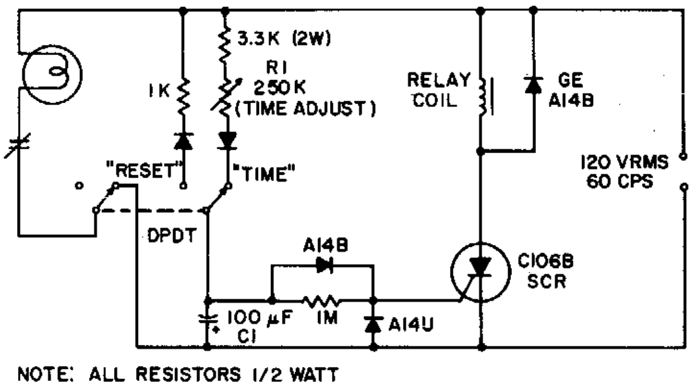


Fig. 88-5

Circuit Notes

This simple timing circuit can delay an output switching function from .01 seconds to about 1 minute. The SCR is triggered by only a few microamps from the timing network R1-C1 to energize the output relay.

0.1 TO 90 SECOND TIMER

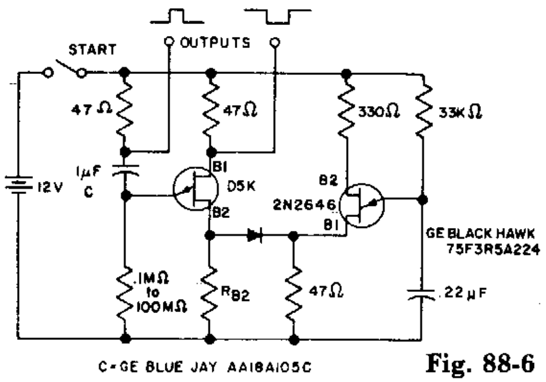


Fig. 88-6

Circuit Notes

The timer interval starts when power is applied to circuit and terminates when voltage is applied to load. 2N2646 is used in oscillator which pulses base 2 of D5K. This reduces the effective I_P of D5K and allows a much larger timing resistor and smaller timing capacitor to be used than would otherwise be possible.

SEQUENTIAL TIMING

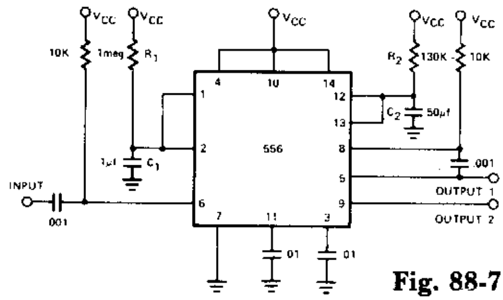


Fig. 88-7

Circuit Notes

By utilizing both halves of the dual timer it is possible to obtain sequential timing. By connecting the output of the first half to the input of the second half via a .001 μF coupling capacitor, sequential timing may be obtained. Delay t_1 is determined by the first half and t_2 by the second half delay. The first half of the timer is started by momentarily connecting pin 6 to ground. When it is timed out (determined by 1.1R1C1) the second half begins. Its time duration is determined by 1.1R2C2.

SOLID-STATE TIMER FOR INDUSTRIAL APPLICATIONS

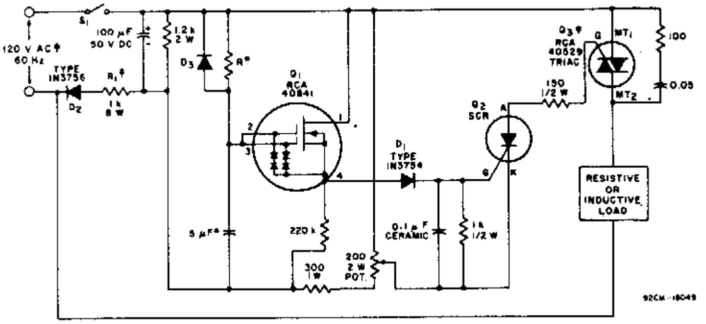


Fig. 88-8

* Cornell-Dubilier Electronics—Type MMW or equivalent.

* R controls duration of time delay. At R = 60 MΩ up to 5-minute delay (IRC resistor, Type CGH or equivalent)

‡ This circuit can also be used at supply voltages of 240 V AC and 24V AC (60Hz) by changing the values of R1 and Q3.

TIMING CIRCUIT CHARACTERISTICS

T_A = −25°C to +60°C

Accuracy: ±10% (over temperature)

Repeatability: ±3% (at 25°C)

Reset Time: Less than 150 ms

Q2: V_{DRM} = 60V
I_{GT} = 200μA
I_T = 0.8A

D3: I_R = 1 nA
V_R = 60V

PRECISION SOLID STATE TIME DELAY CIRCUIT

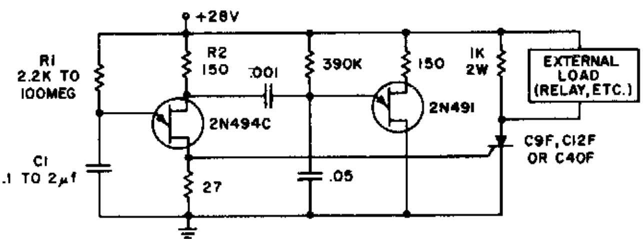
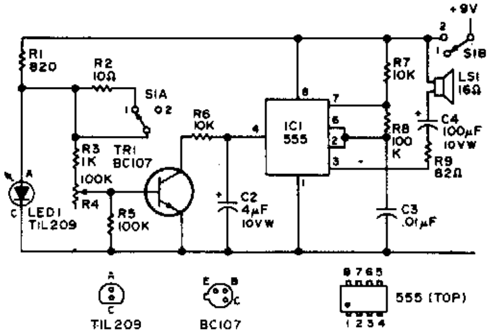


Fig. 88-9

Circuit Notes

Time delays from 0.3 milliseconds to over three minutes are possible with this circuit without using a tantalum or electrolytic capacitor. The timing interval is initiated by applying power to the circuit. At the end of the timing interval, which is determined by the value of R1C1, the 2N494C fires the controlled rectifier. This places the supply voltage minus about one volt across the load. Load currents are limited only by the rating of the controlled rectifier which is from 1 ampere up to 25 amperes for the types specified in the circuit. A calibrated potentiometer could be used in place of R1 to permit setting a predetermined time delay after one initial calibration.

ELECTRONIC EGG TIMER



Circut Notes

The IC functions as an af multivibrator which is controlled by the external transistor. S1A/B is the on-off toggle switch.

Fig. 88-10

ON/OFF CONTROLLER

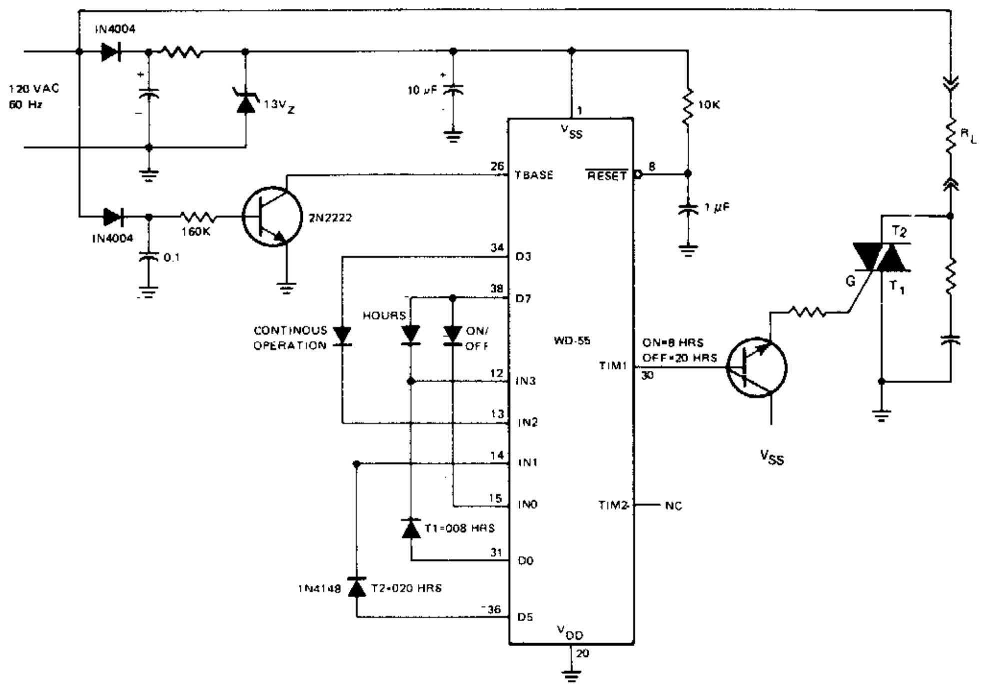
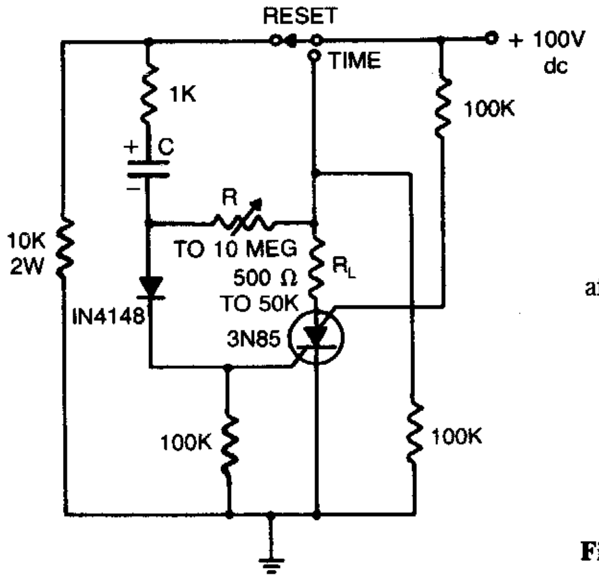


Fig. 88-11

Circuit Notes

The ac line-operated on/off controller is a simple, reliable solid-state alternative to a motive driven cam switch. Time 1 and time 2 are programmed by diodes to be 8 hours and 20 hours respectively. The TIM1 output is buffered by a transistor to supply gate current to a triac which switches the output load. When power is applied to the circuit, the output load is switched on for 8 hours then off for 20 hours repeatedly.

TIMING CIRCUIT



Circuit Notes

Load current starts approximately 0.5 RC after the switch is thrown.

Fig. 88-12

SIMPLE TIMER

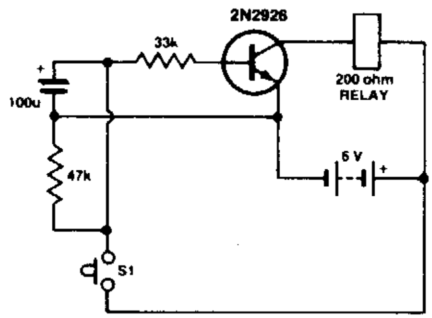
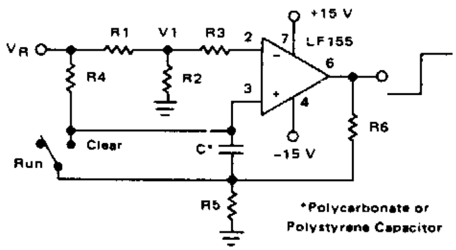


Fig. 88-13

Circuit Notes

Press S1. The 100 μF electrolytic capacitor rapidly charges up at about 0.7 V. The transistor will be forward biased, and collector current will flow operating the relay. Release S1. The capacitor will begin to discharge via the 33 K resistor at the base of the transistor. When the voltage across the capacitor gets down to half a volt or so, the transistor base will no longer be forward biased, collector current will cease, and the relay will drop out. The capacitor will continue to discharge via the 47 K resistor. With the values shown, the relay will remain operated for about eight seconds. Long times are possible with lower values of capacitance by substituting a Darlington pair for the 2N2926. In this case, increase the two resitor values into the megohm range.

LONG INTERVAL RC TIMER



$\text{Time } (t) = R4C\ln(V_R/V_R - V_I)$, $R_3 = R_4$, $R_5 = 0.1\ R6$
If $R1 = R2$: $t = 0.693\ R4C$

Design Example: 100 Second Timer

$V_R = 10$ V	$C = 1\ \mu F$	$R3 = R4 = 144$ M
$R6 = 20$ k	$R5 = 2$ k	$R1 = R2 = 1$ k

Fig. 88-14

741 TIMER

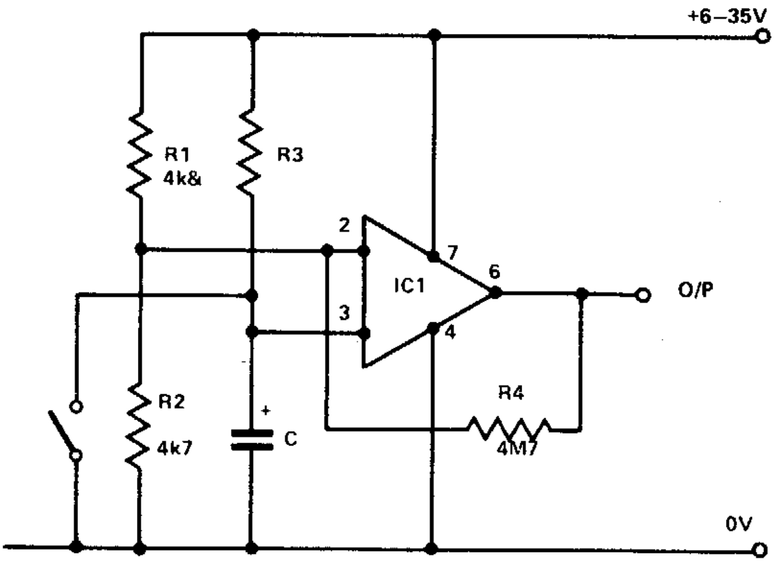


Fig. 88-15

Circuit Notes

R1 and R2 hold the inverting input at half supply voltage. R4 applies feedback to increase the input impedance at pin 3. Pin 3, the noninverting input, is connected to the junction of R3 and C. After the switch is opened, C charges via R3. When the capacitor has charged sufficiently for the potential at pin 3 to exceed that at pin 2 the output abruptly changes from 0 V to positive line potential. If reverse polarity operation is required, simply transpose R3 and C. R3 and C can be any values. Time delays from a fraction of a second to several hours can be obtained by judicious selection. The time delay—independent of supply voltage—is 0.7CR seconds where C is in farads.

TIMER

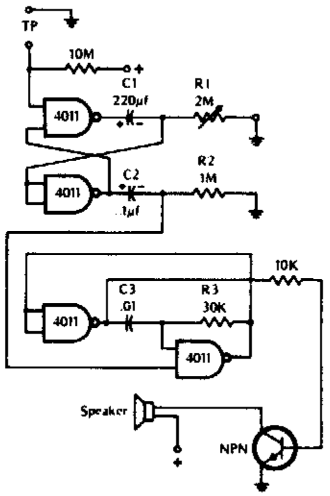


Fig. 88-16

Circuit Notes

The timer can be used wherever time periods of up to seven minutes duration are needed. To turn on just touch the turn-on plate, and after the selected time has elapsed, an alarm will sound for a short period, then automatically turn off. The turn-on touch plate, labeled TP in the diagram, is made up of two metal strips about 1/16-inch apart. Bridging the gap with your finger activates the timer. For more time range, increase R1 and/or C1. R2 and C2 determine the period of time that the alarm will sound. Increasing either will extend the time. The tone of the alarm is determined by R3 and C3. Increasing either lowers the tone, decreasing them raises the tone.

WASHER TIMER

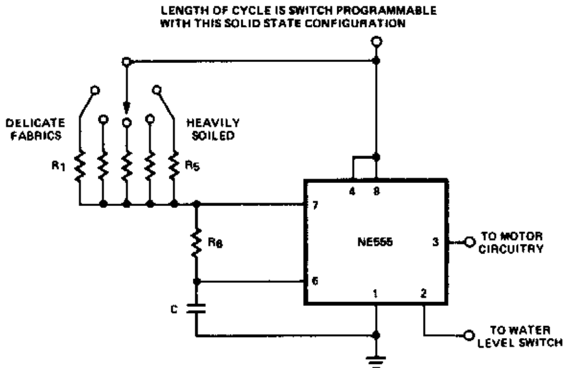


Fig. 88-17

SIMPLE TIME DELAY

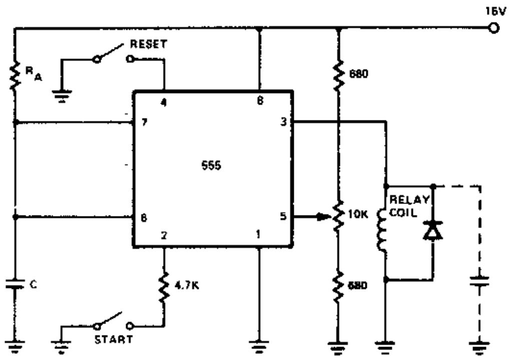


Fig. 88-18

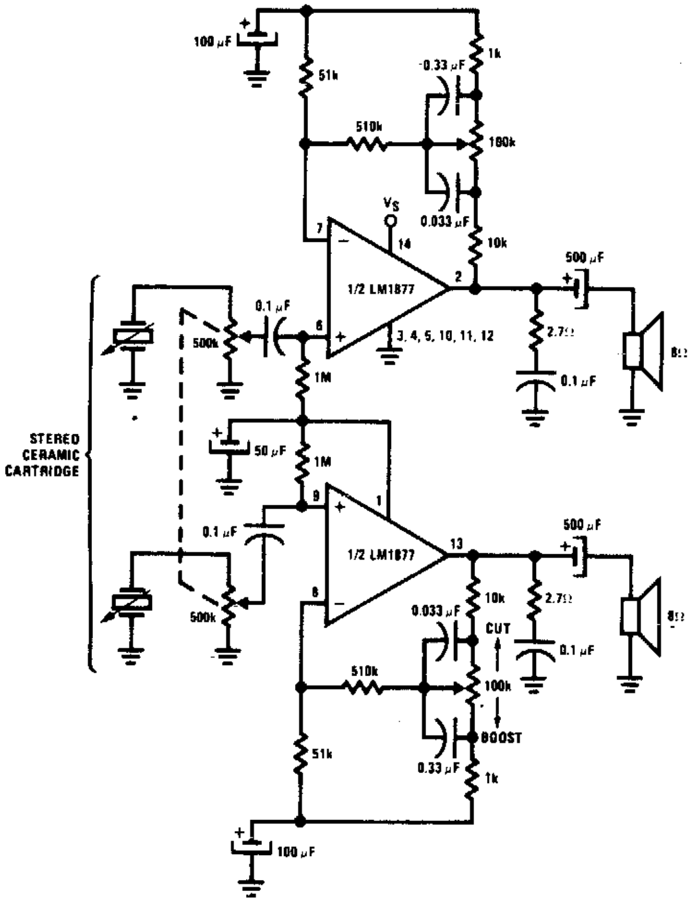
89

Tone Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Stereo Phonograph Amplifier with Bass Tone Control
- Equalizer
- Three-Channel Tone Control
- IC Preamplifier with Tone Control
- Amplifier with Bass Boost
- Active Bass and Treble Tone Control with Buffer
- Passive Bass and Treble Tone Control
- Baxendall Tone-Control Circuit
- High Quality Tone Control
- Microphone Preamplifier with Tone Control
- Hi-Fi Tone Control Circuit
- Three-Band Active Tone Control
- Tone Control Circuit

STEREO PHONOGRAPH AMPLIFIER WITH BASS TONE CONTROL



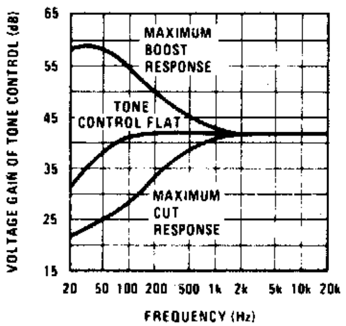
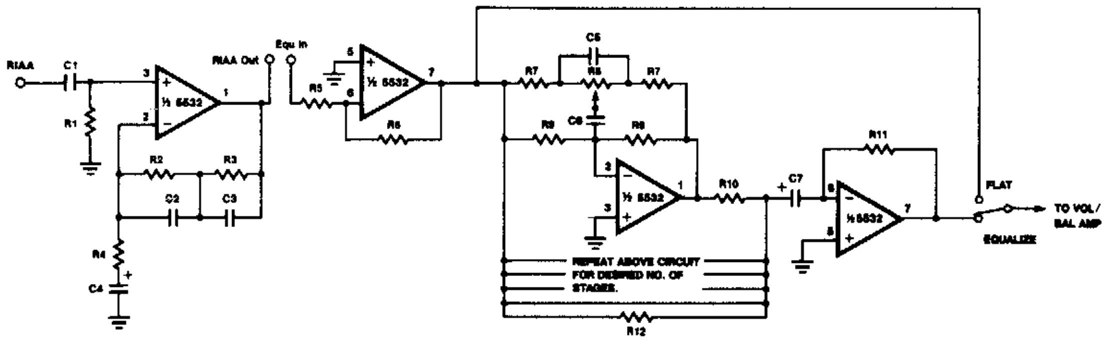


Fig. 89-1

EQUALIZER



COMPONENT VALUE TABLES

R8 = 25k R7 = 2.4k R9 = 240k			R8 = 50k R7 = 5.1k R9 = 510k			R8 = 100k R7 = 10k R9 = 1meg		
fo	C5	C6	fo	C5	C6	fo	C5	C6
23 Hz	1µF	.1µF	25 Hz	.47µF	.047µF	12 Hz	.47µF	.047µF
50 Hz	.47µF	.047µF	36 Hz	.33µF	.033µF	18 Hz	.33µF	.033µF
72 Hz	.33µF	.033µF	54 Hz	.22µF	.022µF	27 Hz	.22µF	.022µF
108 Hz	.22µF	.022µF	79 Hz	.15µF	.015µF	39 Hz	.15µF	.015µF
158 Hz	.15µF	.015µF	119 Hz	.1µF	.01µF	59 Hz	.1µF	.01µF
238 Hz	.1µF	.01µF	145 Hz	.082µF	.0082µF	72 Hz	.082µF	.0082µF
290 Hz	.082µF	.0082µF	175 Hz	.068µF	.0068µF	87 Hz	.068µF	.0068µF
350 Hz	.068µF	.0068µF	212 Hz	.056µF	.0056µF	106 Hz	.056µF	.0056µF
425 Hz	.056µF	.0056µF	253 Hz	.047µF	.0047µF	126 Hz	.047µF	.0047µF
506 Hz	.047µF	.0047µF	360 Hz	.033µF	.0033µF	180 Hz	.033µF	.0033µF
721 Hz	.033µF	.0033µF	541 Hz	.022µF	.0022µF	270 Hz	.022µF	.0022µF
1082 Hz	.022µF	.0022µF	794 Hz	.015µF	.0015µF	397 Hz	.015µF	.0015µF
1588 Hz	.015µF	.0015µF	1191 Hz	.01µF	.001µF	595 Hz	.01µF	.001µF
2382 Hz	.01µF	.001µF	1452 Hz	.0082µF	820pF	726 Hz	.0082µF	820pF
2904 Hz	.0082µF	820pF	1751 Hz	.0068µF	680pF	875 Hz	.0068µF	680pF
3502 Hz	.0068µF	680pF	2126 Hz	.0056µF	560pF	1063 Hz	.0056µF	560pF
4253 Hz	.0056µF	560pF	2534 Hz	.0047µF	470pF	1267 Hz	.0047µF	470pF
5068 Hz	.0047µF	470pF	3609 Hz	.0033µF	330pF	1804 Hz	.0033µF	330pF
7218 Hz	.0033µF	330pF	5413 Hz	.0022µF	220pF	2706 Hz	.0022µF	220pF
10827 Hz	.0022µF	220pF	7940 Hz	.0015µF	150pF	3970 Hz	.0015µF	150pF
15880 Hz	.0015µF	150pF	11910 Hz	.001µF	100pF	5955 Hz	.001µF	100pF
23820 Hz	.001µF	100pF	14524 Hz	820pF	82pF	7262 Hz	820pF	82pF
			17514 Hz	680pF	68pF	8757 Hz	680pF	68pF
			21267 Hz	560pF	56pF	10633 Hz	560pF	56pF
						12670 Hz	470pF	47pF
						18045 Hz	330pF	33pF

COMPONENT VALUES

R1	1meg	C1	.22µF
R2	100k	C2	750pF
R3	1meg	C3	.0033µF
R4	1.1k	C4	33µF
R5	100k	C5	SEE TABLE
R6	100k	C6	SEE TABLE
R7	SEE TABLE	C7	2.2µF
R8	(pot) SEE TABLE		
R9	SEE TABLE		
R10	100k		
R11	100k		
R12	20k (5 STAGES)		

Fig. 89-2

THREE-CHANNEL TONE CONTROL

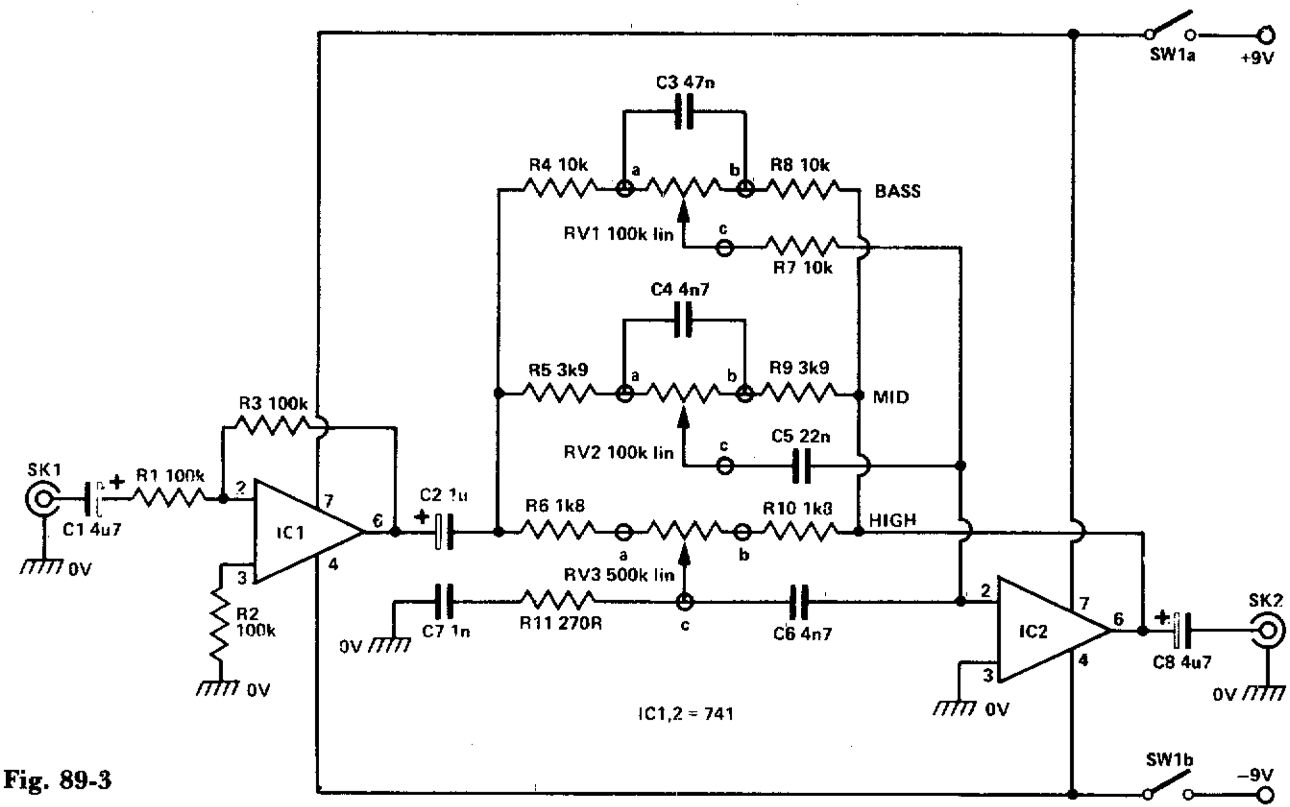


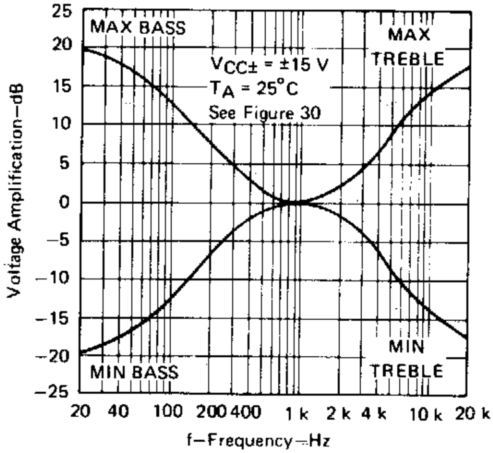
Fig. 89-3

Circuit Notes

The input signal is fed via SK1 to the first active stage built around IC1. Configured as a noninverting amplifier whose gain is set by the ratio of R3 and R1. In this case, the gain is set at unity. This initial stage is required to isolate the following stage from any loading effects. The output from IC1 is fed via three frequency shaping networks to IC2. The three networks built around RV1, RV2, and RV3 are also included in the feedback path of IC2, another inverting op amp stage. The components associated with the three variable resistors are chosen to give the required frequency control.

IC PREAMPLIFIER WITH TONE CONTROL

IC PREAMPLIFIER RESPONSE CHARACTERISTICS



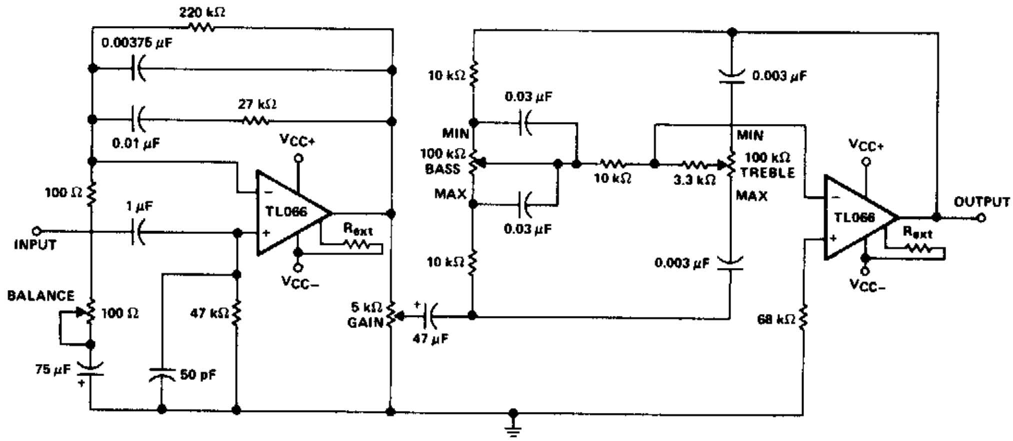
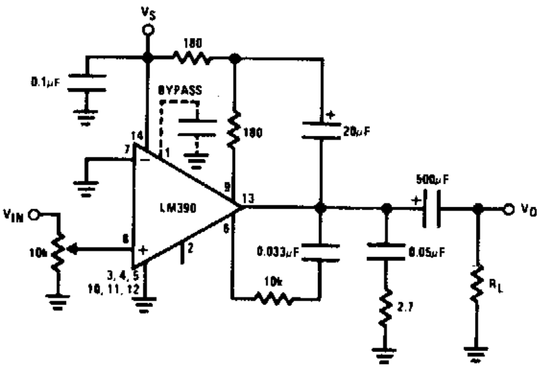
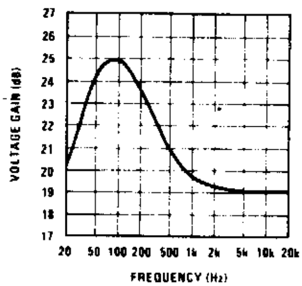


Fig. 89-4

AMPLIFIER WITH BASS BOOST





Frequency Response with Bass Boost

Fig. 89-5

ACTIVE BASS & TREBLE TONE CONTROL WITH BUFFER

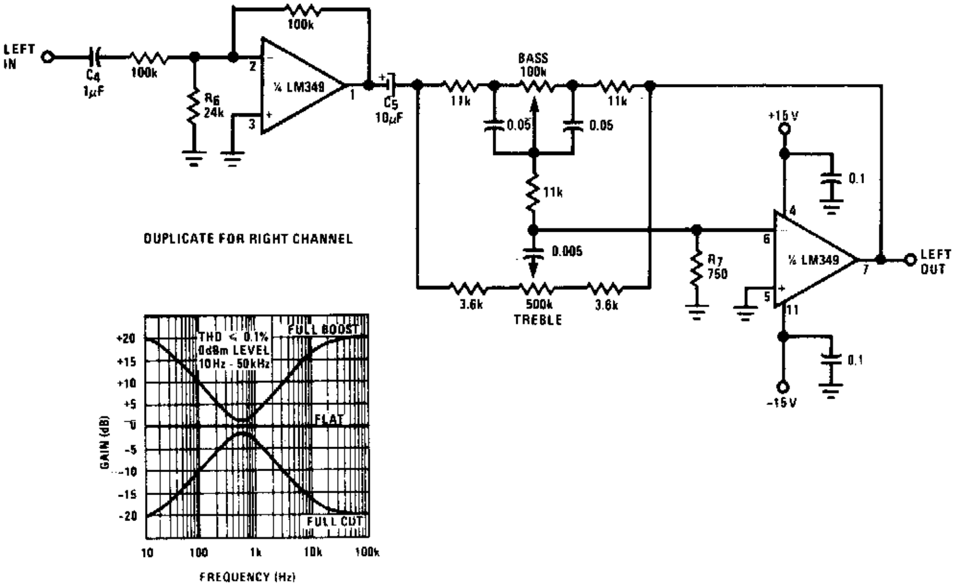
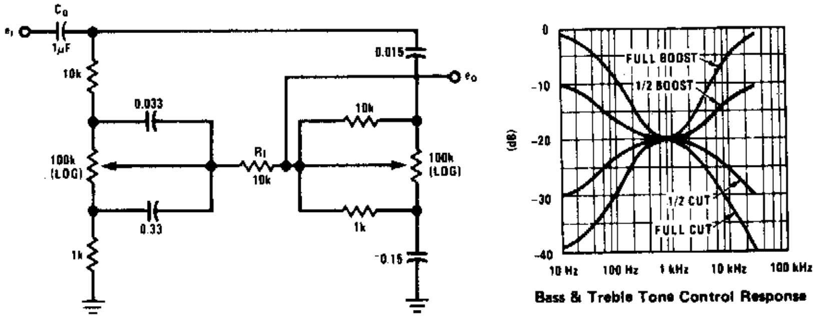


Fig. 89-6

PASSIVE BASS & TREBLE TONE CONTROL



Bass & Treble Tone Control Response

Fig. 89-7

BAXENDALL TONE-CONTROL CIRCUIT

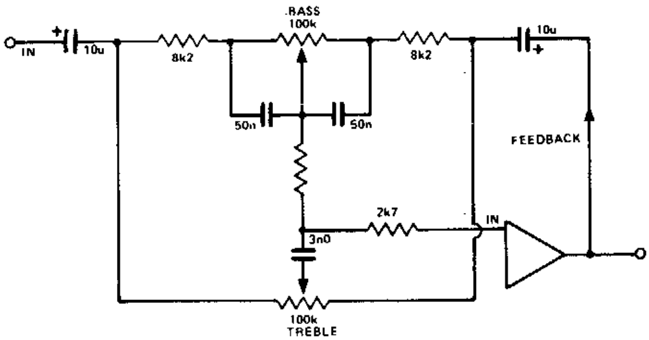


Fig. 89-8

HIGH QUALITY TONE CONTROL

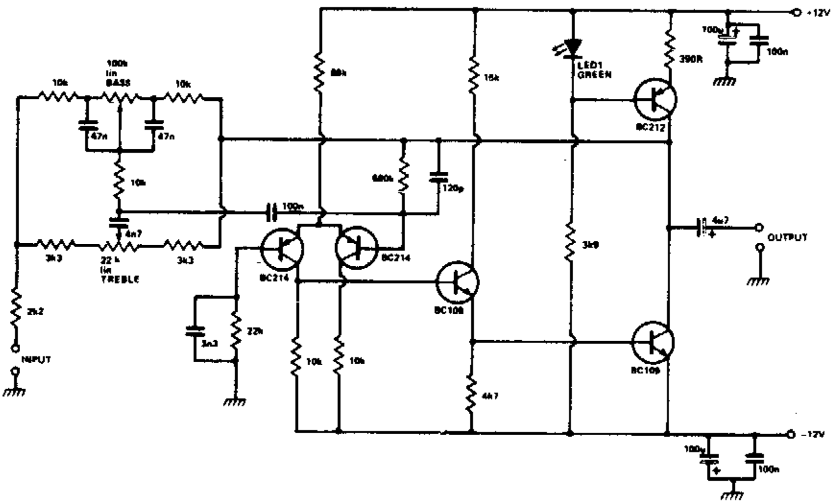


Fig. 89-9

Circuit Notes

The circuit is based on an inverting op amp using discrete transistors to overcome poor slew rate, fairly high distortion, and high noise problems. The output stage is driven by a constant current source, biased by a green LED to provide temperature compensation. With the controls flat, the unit provides unity gain so the stage can be switched in or out. The design is suitable for inputs between 100 mV and 1 V and provides a good overload margin at low distortion for the accurate reproduction of transients. The usual screening precautions against hum should be carried out.

MICROPHONE PREAMPLIFIER WITH TONE CONTROL

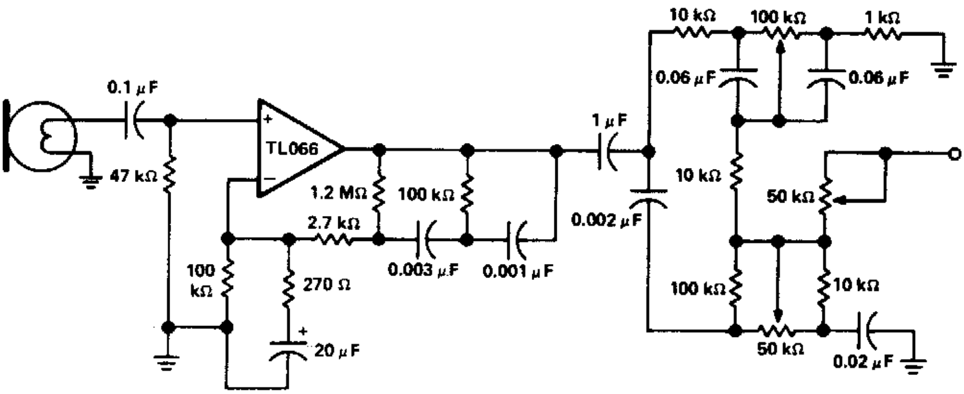


Fig. 89-10

HI-FI TONE CONTROL CIRCUIT (HIGH Z INPUT)

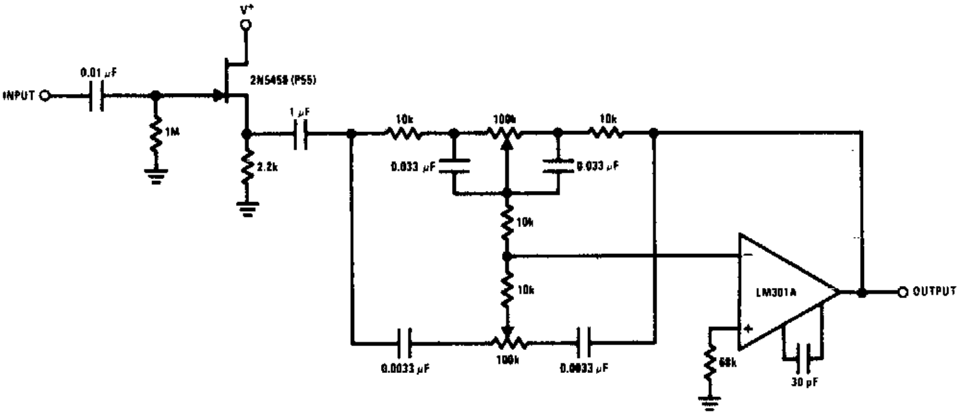


Fig. 89-11

Circuit Notes

The 2N5458 JFET provides the function of a high input impedance and low noise characteristics to buffer an op amp feedback tone control circuit.

THREE-BAND ACTIVE TONE CONTROL

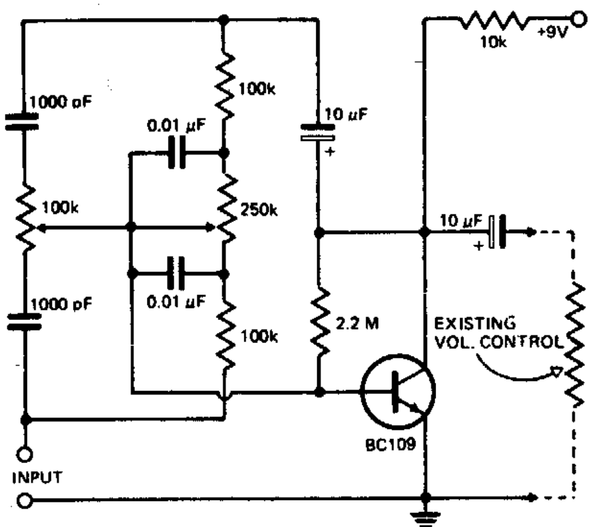
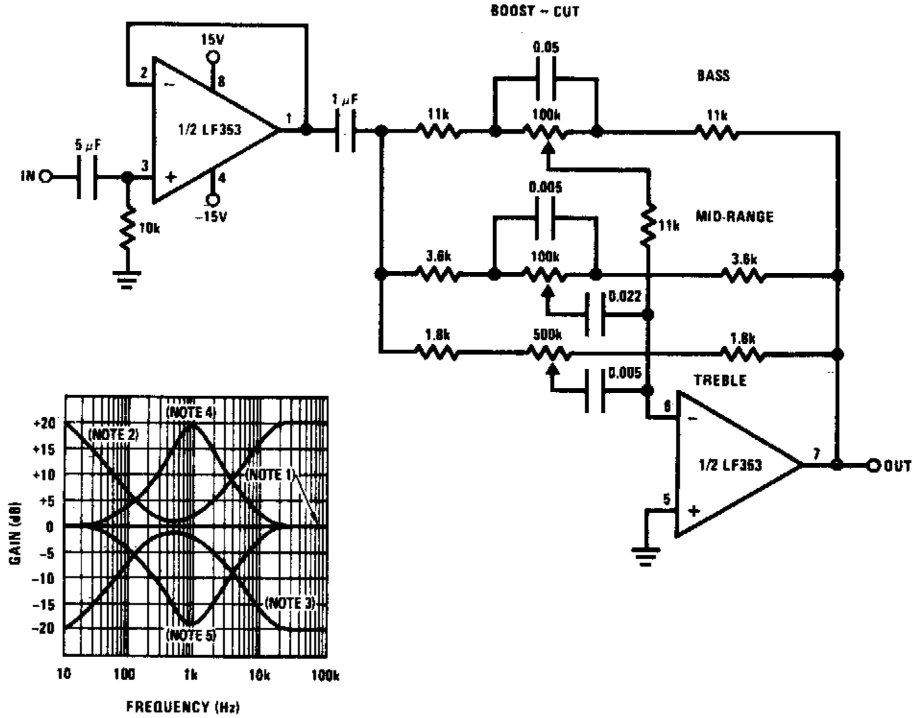


Fig. 89-12

TONE CONTROL CIRCUIT



Note 1: All controls flat.
Note 2: Bass and treble boost, mid flat.
Note 3: Bass and treble cut, mid flat.
Note 4: Mid boost, bass and treble flat.
Note 5: Mid cut, bass and treble flat.

- All potentiometers are linear taper
- Use the LF347 Quad for stereo applications

Fig. 89-13

Circuit Notes

A simple single-transistor circuit will give approximately 15 dB boost or cut at 100 Hz and 15 kHz respectively. A low noise audio type transistor is used, and the output can be fed directly into any existing amplifier volume control to which the tone control is to be fitted. The gain of the circuit is near unity when controls are set in the flat position.

90

Transmitters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

WIRELESS AM MICROPHONE

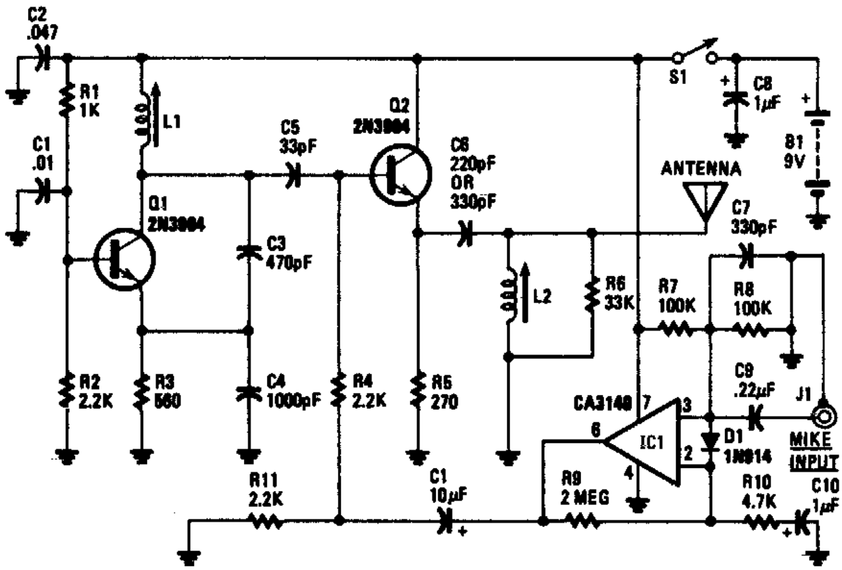


Fig. 90-1

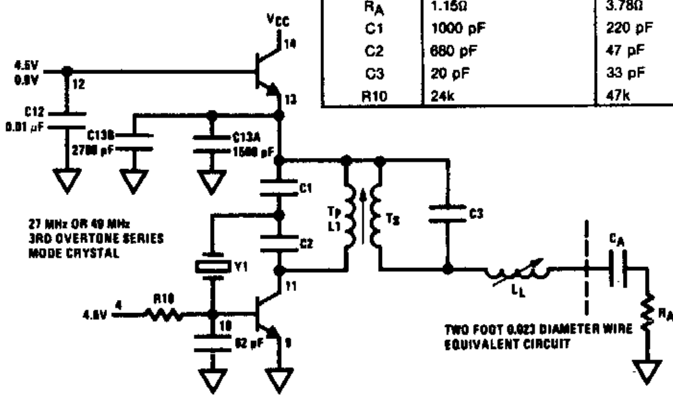
Circuit Notes

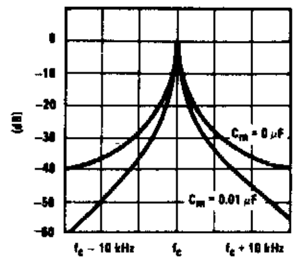
Transistor Q1 and its associated components comprise a tuneable rf oscillator. The rf signal is fed to transistor Q2, the modulator. Operational amplifier IC1 increases the audio signal and applies it through resistor R4 to the base of Q2. Tune an AM radio to an unused frequency between 800 to 1600 kHz. Tune L1 for a change in the audio level coming from the radio. Peak the output by adjusting L2. If L1 is disturbed, it may be necessary to readjust L2 for peak performance. Depending on the impedance of the microphone audio sensitivity can be increased by decreasing the value of R10 and vice versa.

27 MHz AND 49 MHz RF OSCILLATOR/TRANSMITTER

Component	27 MHz	49 MHz
T_P	2 Turns	6 Turns
T_S	3 Turns	1 Turn
L1	TOKO KXN K4636 BJF	TOKO KEN K4635 BJE
L_L	MILLER #4611	MILLER #9338-10
C_A	5.4 pF	6.2 pF
R_A	1.15Ω	3.78Ω
C1	1000 pF	220 pF
C2	680 pF	47 pF
C3	20 pF	33 pF
R10	24k	47k

Use TOKO form #51-0116-02 and #30 wir or #51-0178 and #32 wire





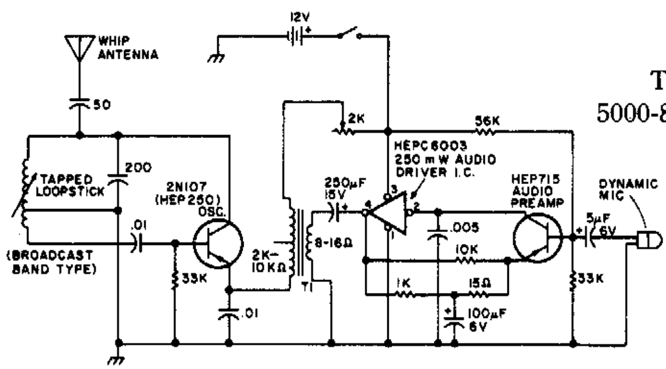
Envelope of Transmitted Spectrum for Circuit

Fig. 90-2

Circuit Notes

The modulator and oscillator consist of two NPN transistors. The base of the modulator transistor is driven by a bidirectional current source with the voltage range for the high condition limited by a saturating PNP collector to the pin 4 V_{REG} voltage and low condition limited by a saturating NPN collector in series with a diode to ground. The crystal oscillator/transmitter transistor is configured to oscillate in a class C mode. Because third overtone crystals are used for 27 MHz or 49 MHz applications a tuned collector load must be used to guarantee operation at the correct frequency.

1-2 MHz BROADCAST TRANSMITTER



Circuit Notes

T1 is a low impedance output transformer 5000-8 ohms.

Fig. 90-3

ONE TUBE, 10 WATT C.W. TRANSMITTER

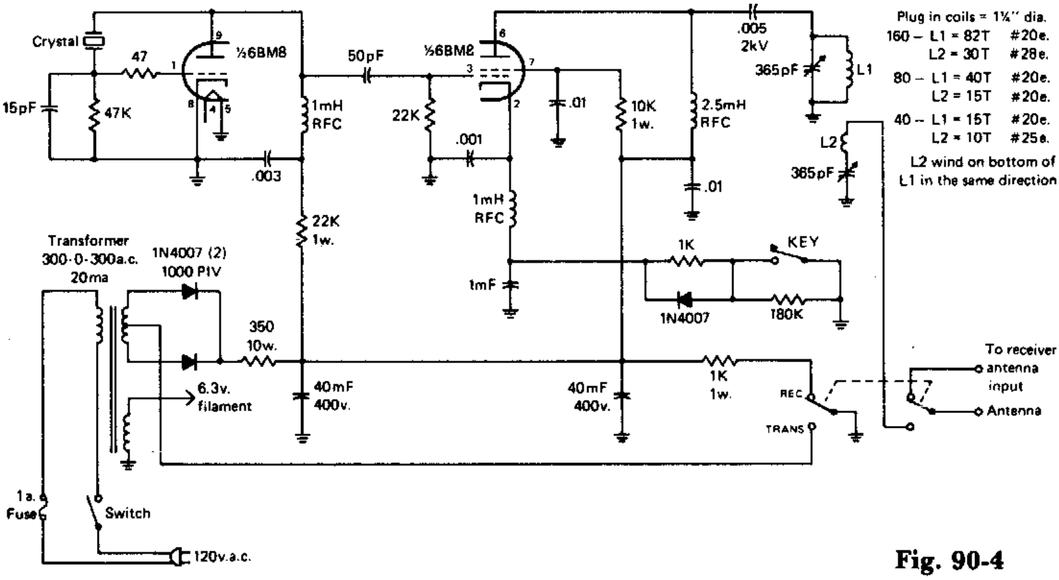


Fig. 90-4

SIMPLE FM TRANSMITTER

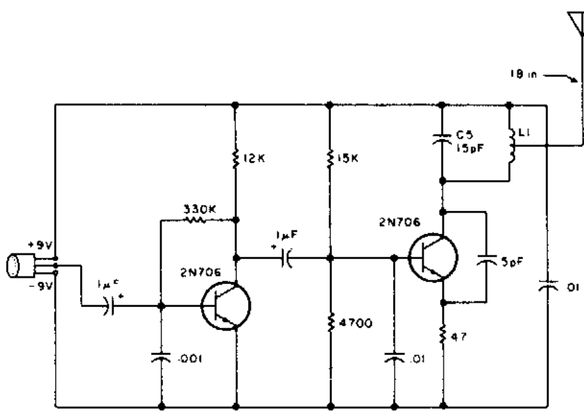


Fig. 90-5

Circuit Notes

This transmitter can be tuned to the FM broadcast band, 2 meters, or other VHF bands by changing C5 and L1. The values given for C5 and L1 will place the frequency somewhere in the FM broadcast band. L1 is 4 turns of #20 enameled wire airwound, ¼ inch in diameter, 5mm long and center-tapped. The microphone is an electret type and the antenna is 18 inches of any type of wire. Keep all leads as short as possible to minimize stray capacitance. The range of the transmitter is several hundred yards.

91

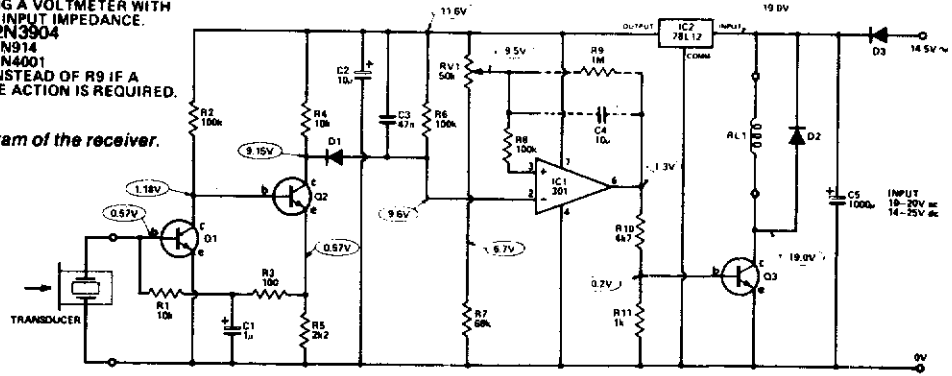
Ultrasonics

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ultrasonic Switch
Ultrasonic Bug-Chaser
Ultrasonic Pest Repeller
Mosquito-Repelling Circuit
40 kHz Ultrasonic Transmitter

ULTRASONIC SWITCH

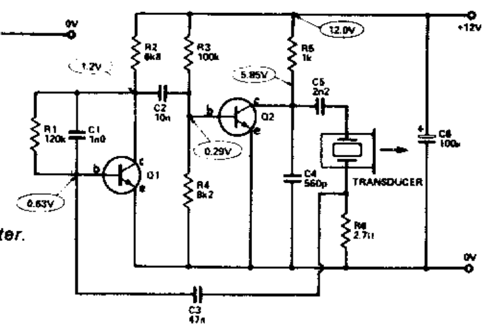
NOTES:
VOLTAGES MEASURED WITH NO INPUT SIGNAL USING A VOLTMETER WITH 10 MEG OHM INPUT IMPEDANCE.
Q1–Q3 ARE 2N3904
D1 IS 1N914
D2,D3 ARE 1N4001
C4 IS USED INSTEAD OF R9 IF A MONOSTABLE ACTION IS REQUIRED.



Circuit diagram of the receiver.

NOTE:
VOLTAGES MEASURED USING A VOLTMETER WITH 10 MEG OHM INPUT IMPEDANCE.

Q1-2 ARE 2N3904



Circuit diagram of the transmitter.

Fig. 91-1

Circuit Notes

Receiver. Output from the transducer is amplified by Q1 and Q2, and rectified by D1. Voltage on pin 2 of IC1 will go more negative as the input signal increases. IC1 is used as a comparator and checks the voltage on pin 2 (i.e., the sound level), to that on pin 3 which is the reference level. If pin 2 is at a lower voltage than pin 3 (i.e., a signal is present), the output of IC1 will be high (about 10.5 volts) and this will turn on Q3 which will close the relay. The converse occurs if pin 2 is at a higher voltage than pine 3.

Transmitter. The oscillator frequency is determined by the transducer characteristics [(minimum (series resonance) at 39.8 kHz followed by a maximum (parallel resonance) at 41.5 kHz.)] Two transistors from a noninverting amplifier and positive feedback is supplied via the transducer, R6 and C3. At the series resonant frequency, this feedback is strong enough to cause oscillation.

ULTRASONIC BUG-CHASER

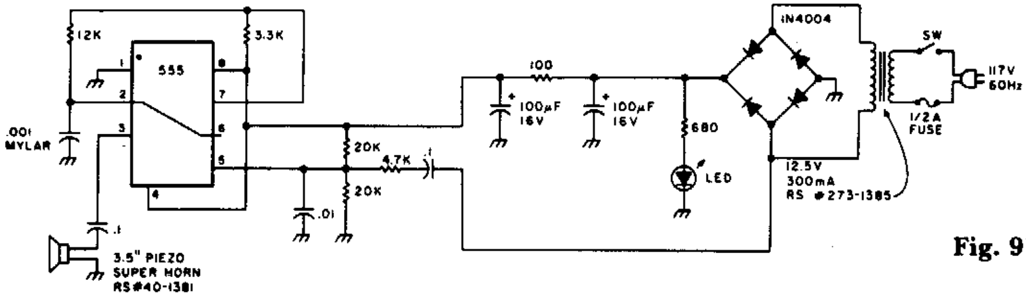


Fig. 91-2

Circuit Notes

Low-intensity ultrasonic sound waves in the 30-45 kHz frequency band repel insects and small rodents. The unit is designed to generate a swept square wave from 30 to 45 kHz. The LM555 IC is wired as an ultrasonic oscillator driving a piezoelectric speaker of the hi-fi super-tweeter type. The output of the oscillator is swept by a 60-Hz signal from the ac input of the bridge rectifier. The LED acts as a pilot.

MOSQUITO-REPELLING CIRCUIT

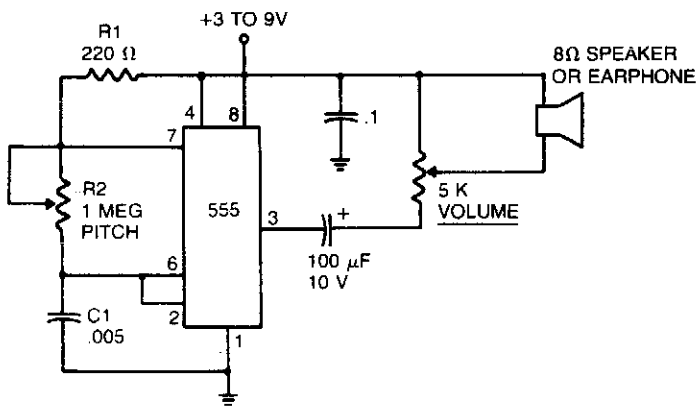


Fig. 91-3

Circuit Notes

In the 555 oscillator circuit, adjusting R2 will provide output frequencies from below 200 Hz to above 62 kHz. Use a good quality miniature speaker so that it will produce frequencies on the order of 20 kHz.

ULTRASONIC PEST REPELLER

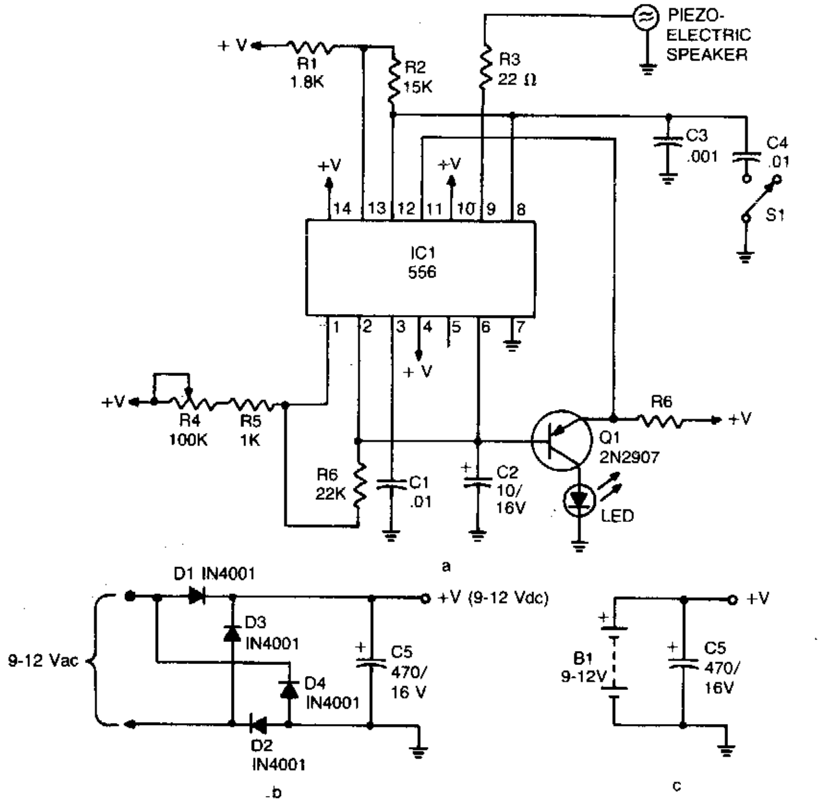


Fig. 91-4

Circuit Notes

The device emits ultrasonic sound waves that sweep between 65,000 and 25,000 hertz. Designed around a 556 dual timer, one half operated as an astable multivibrator with an adjustable frequency of 1 to 3 Hz. The second half is also operated as an astable multivibrator but with a fixed free running frequency around 45,000 Hz. The 25-65 kHz sweep is accomplished by coupling the voltage across C2 (the timing capacitor for the first half of the 556) via Q1 to the control voltage terminal (pin 11) of the second half of the 556. The device that radiates the ultrasonic sound is a piezo tweeter.

40 kHz ULTRASONIC TRANSMITTER

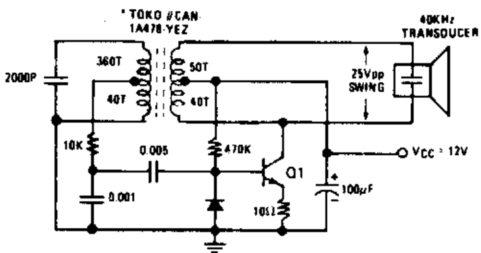


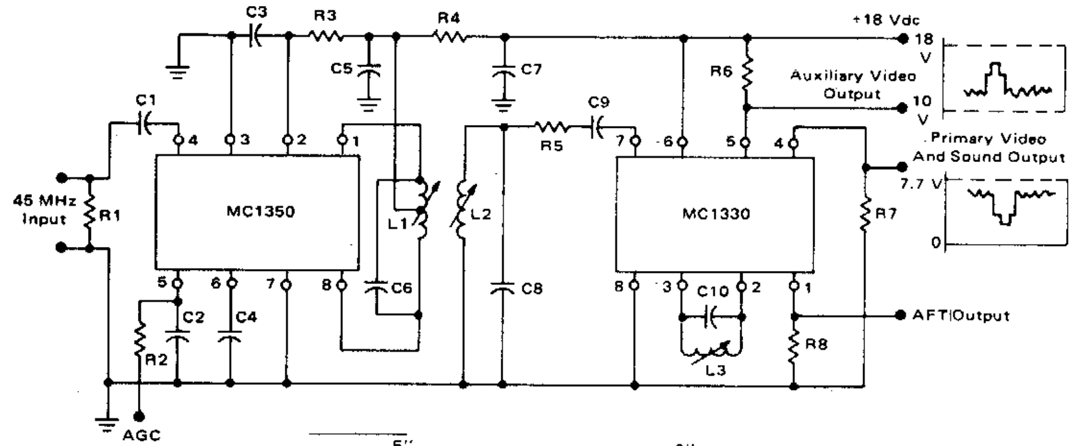
Fig. 91-5

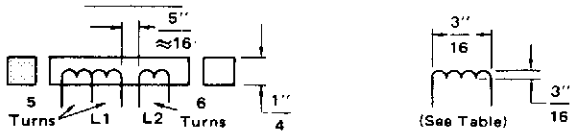
92

Video Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VIDEO IF AMPLIFIER AND LOW-LEVEL VIDEO DETECTOR CIRCUIT





All windings #30 AWG tinned nylon acetate wire tuned with high permiability slugs. Coil Craft #4786 differential transformer.

L3 wound with #26 AWG tinned nylon acetate wire tuned by distorting winding.

C1 = 0.001 μF
C2 = 0.002 μF
C3 = 0.002 μF
C4 = 0.002 μF
C5 = 0.002 μF
All Caps Marked μF Ceramic HiK
All Caps Marked pF Silver Mica 5%

C6 = See Table
C7 = 0.1 μF
C8 = See Table
C9 = 68 pF
C10 = See Table

R1 = 50 Ω
R2 = 5 k
R3 = 470 Ω
R4 = 220 Ω
R5 = 22 Ω

R6 = 3.3 kΩ
R7 = 3.9 kΩ
R8 = 3.9 kΩ
All Resistors 1/4-W ± 10%

Table of Component Values

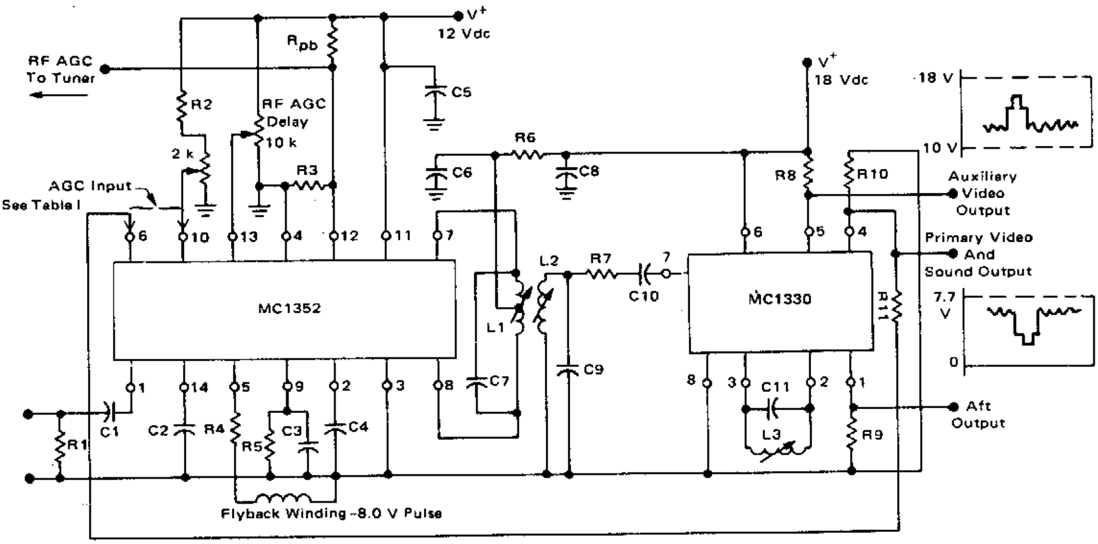
Component	36 MHz	45 MHz	58 MHz
C6	24 pF	15 pF	10 pF
C8	18 pF	12 pF	10 pF
C10	33 pF	33 pF	18 pF
L3	12 Turns	10 Turns	10 Turns

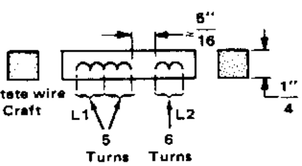
Fig. 92-1

Circuit Notes

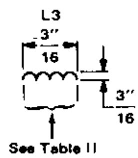
The circuit has a typical voltage gain of 84 dB and a typical AGC range of 80 dB. It gives very small changes in bandpass shape, usually less than 1 dB tilt for 60 dB compression. There are no shielded sections. The detector uses a single tuned circuit (L3 and C10). Coupling between the two integrated circuits is achieved by a double tuned transformer (L1 and L2).

TELEVISION IF AMPLIFIER AND DETECTOR USING AN MC1330 AND AN MC1352





All windings #30 AWG tinned nylon acetate wire tuned with high permiability slugs. Coil Craft #4786 differential transformer.



C10 = 62 pF
C11 = (See Table II)
All Resistors 1/4-Watt ±5%

Wound with #26 AWG tinned nylon acetate wire tuned by distorting winding.

TABLE I

Video Polarity	Pin 6 Voltage	Pin 10 Voltage	R4
Negative-Going Sync.	5.5 2.0 0	Adj. 1.0–4.0 Vdc Nom 2.0 V	0
Positive-Going Sync.	Adj. 1.0–8.0 Vdc Nom 4.5 V	4.5 0	3.9 k

TABLE II

Component	36 MHz	45 MHz	58 MHz
C7	24 pF	15 pF	10 pF
C9	18 pF	12 pF	10 pF
C11	33 pF	33 pF	18 pF
L3	12 Turns	10 Turns	10 Turns

R_{pb} (See Text)
R1 = 50 Ω
R2 = 3.9 kΩ
R3 = (See Text)
R4 = (See Table I)
R5 = 220 kΩ
R6 = 220 Ω
R7 = 22 Ω
R8 = 3.3 kΩ
R9 = 3.9 kΩ
R10 = 3.9 kΩ
R11 = 4.7 kΩ
C1 = 0.001 μF
C2 = 0.1 μF
C3 = 0.25 μF
C5 = 0.1 μF
C6 = 0.1 μF
C7 = (See Table II)
C8 = 0.1 μF
C9 = (See Table II)

Fig. 92-2

TWO-STAGE WIDEBAND AMPLIFIER

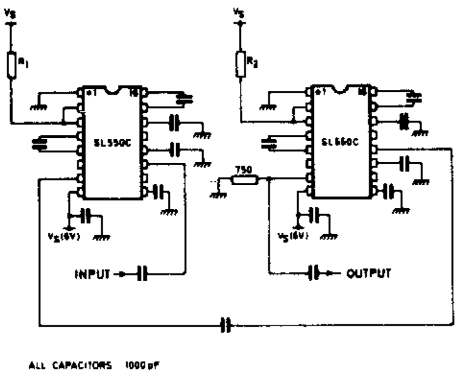
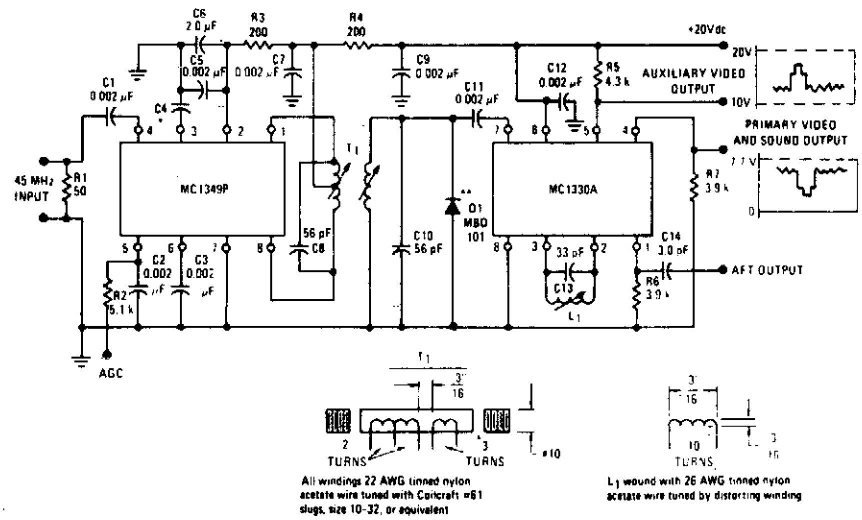


Fig 92-3

Circuit Notes

A wideband high gain configuration using two SL550s connected in series. The first stage is connected in common emitter configuration, the second stage is a common base circuit. Stable gains of up to 65 dB can be achieved by the proper choice of R1 and R2. The bandwidth is 5 to 130 MHz, with a noise figure only marginally greater than the 2.0 dB specified for a single stage circuit.

VIDEO IF AMPLIFIER AND LOW-LEVEL VIDEO DETECTOR CIRCUIT

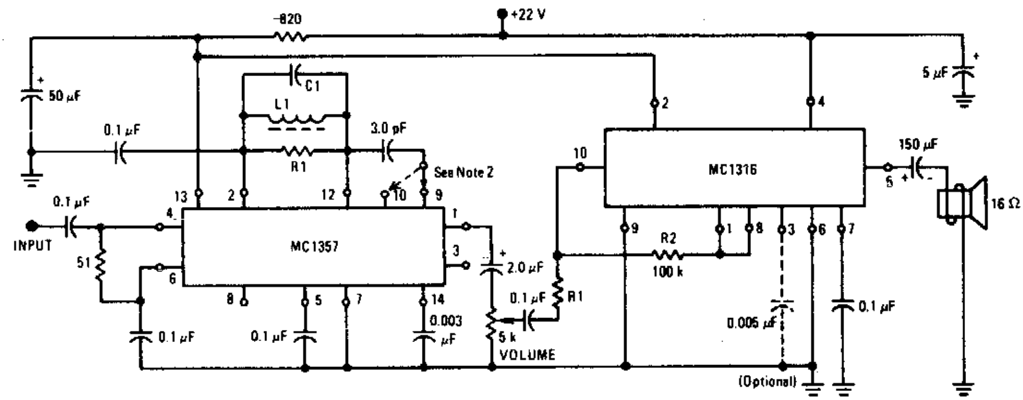


*See Note 1 (page 3), and C4, Parts List (page 4) for this specification on the MC1349P Data Sheet

**See Input Overload Section of the Design Characteristics Page 3, and General Information, Page 5 Note 6

Fig. 92-4

TV SOUND IF OR FM IF AMPLIFIER WITH QUADRATURE DETECTOR



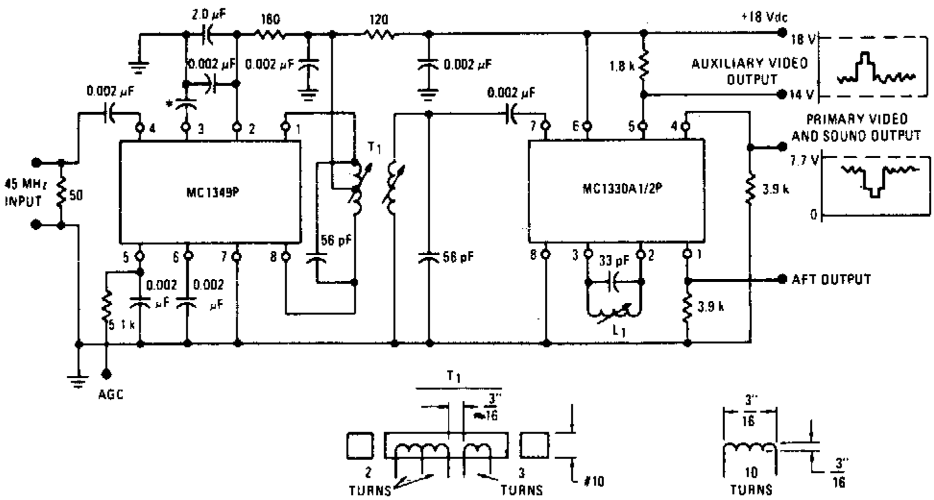
Typical Performance:
2 Watts Output
2% Distortion
250 µV Sensitivity (3 dB Lim.)

C1 = 120 pF
L1 = 14 µH
R1 = 20 kΩ
Q = 30

Fig. 92-5

IF AMPLIFIER

– TYPICAL APPLICATION OF MC1349P VIDEO IF AMPLIFIER and MC1330 LOW-LEVEL VIDEO DETECTOR CIRCUIT



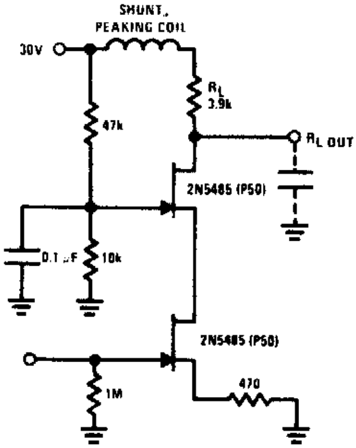
All windings #22 AWG tinned nylon acetate wire tuned with Coilcraft #61 slugs, size 10-32, or equivalent.

L1 wound with #26 AWG tinned nylon acetate wire tuned by distorting winding.

*See Note 1 (page 3), and C4, Parts List (page 4) of this specification.

Fig. 92-6

FET CASCODE VIDEO AMPLIFIER

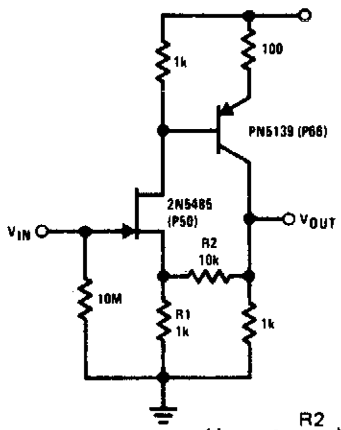


Circuit Notes

The FET cascode video amplifier features very low input loading and reduction of feedback to almost zero. The 2N5485 is used because of its low capacitance and high Y_{fs}. Bandwidth of this amplifier is limited by RL and load capacitance.

Fig. 92-7

HIGH IMPEDANCE LOW CAPACITANCE AMPLIFIER



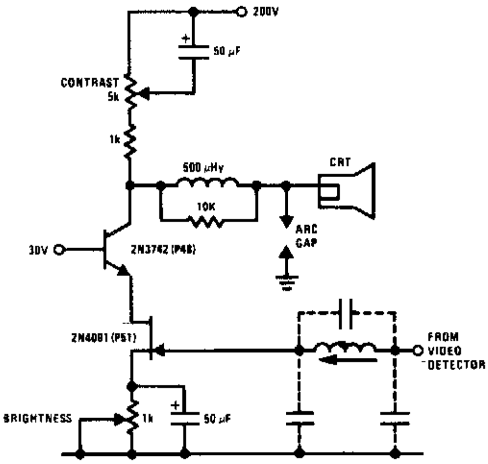
$$V_{OUT} \simeq \frac{R2}{R1} V_{IN}$$

Fig. 92-8

Circuit Notes

This compound series-feedback circuit provides high input impedance and stable, wide-band gain for general purpose video amplifier applications.

JFET BIPOLAR CASCODE VIDEO AMPLIFIER



Circuit Notes

The JFET-bipolar cascode circuit will provide full video output for the CRT cathode drive. Gain is about 90. The cascode configuration eliminates Miller capacitance problems with the 2N4091 JFET, thus allowing direct drive from the video detector. An m-derived filter using stray capacitance and a variable inductor prevents 4.5 MHz sound frequency from being amplified by the video amplifier.

Fig. 92-9

VIDEO AMPLIFIER

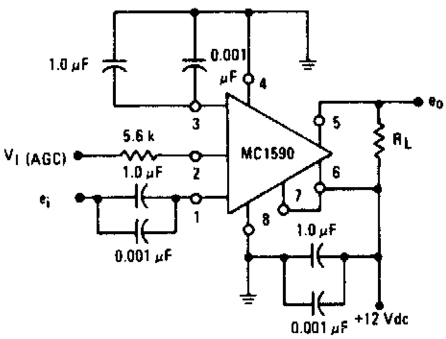
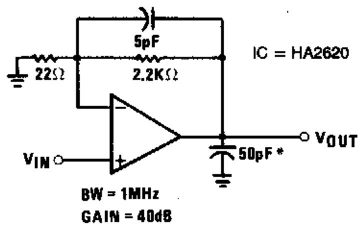


Fig. 92-10

VIDEO AMPLIFIER



*A small load capacitance of at least 30pF (including stray capacitance) is recommended to prevent possible high frequency oscillations.

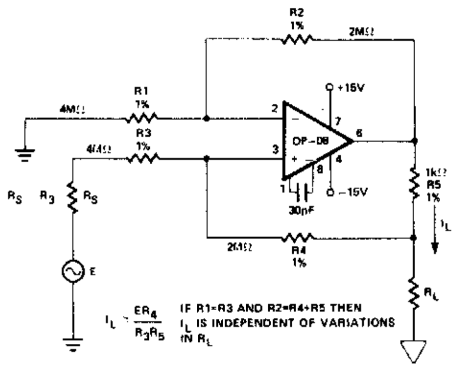
Fig. 92-11

93

Voltage and Current Sources and References

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

BILATERAL CURRENT SOURCE



Circuit Notes

The circuit will produce the current relationship to within 2% using 1% values for R1 through R5. This includes variations in RL from 100 ohm to 2000 ohm. The use of large resistors for R1 through R4 minimizes the error due to RL variations. The large resistors are possible because of the excellent input bias current performance of the OP-08.

Fig. 93-1

0 V TO 20 V POWER REFERENCE

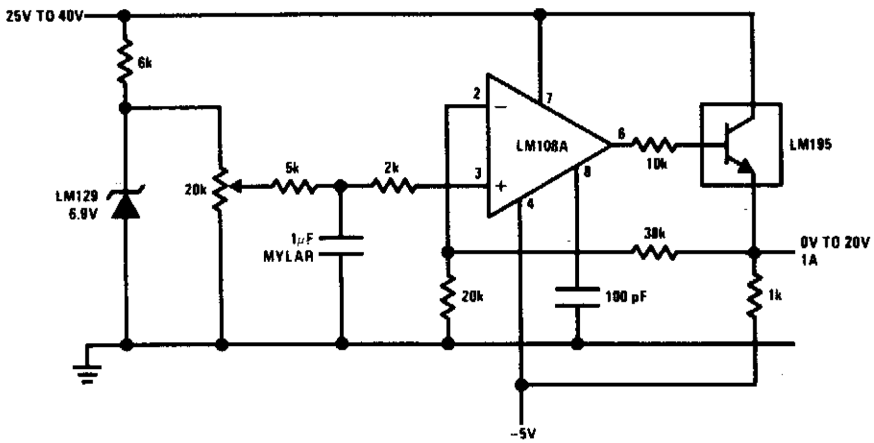


Fig. 93-2

PROGRAMMABLE VOLTAGE SOURCE

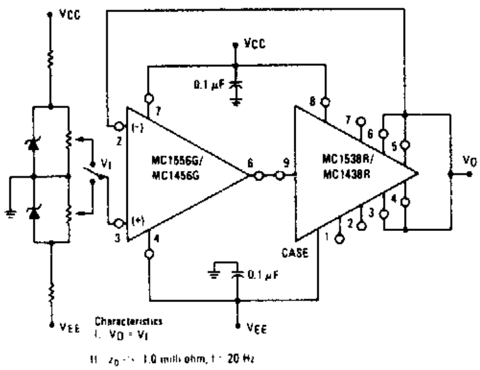


Fig. 93-3

BILATERAL CURRENT SOURCE

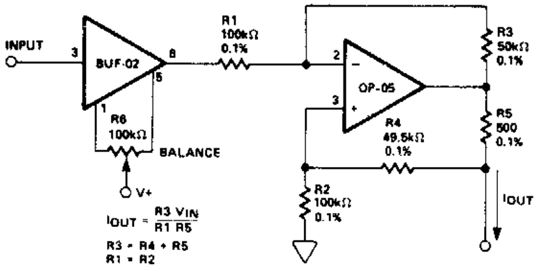


Fig. 93-4

LOW VOLTAGE ADJUSTABLE REFERENCE SUPPLY

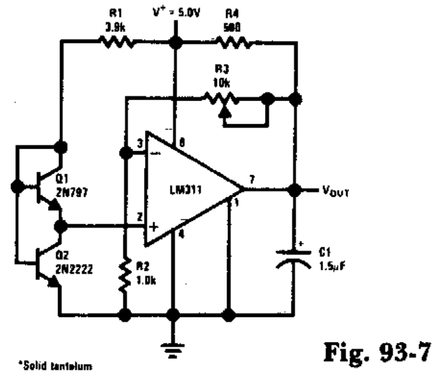


Fig. 93-7

NONINVERTING BIPOLAR CURRENT SOURCE

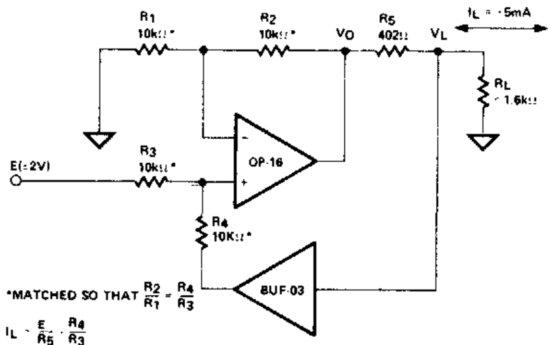


Fig. 93-5

COMPLIANCE OF ABOVE CIRCUIT (ΔV_L) IS ±8V WHEN E = ±2V AND $R_L \le 1.6k\Omega$. NOTE THAT V_O IS ±10V UNDER THESE CONDITIONS.

VOLTAGE REFERENCE

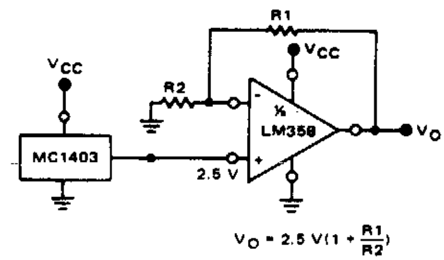


Fig. 93-8

VOLTAGE REFERENCE

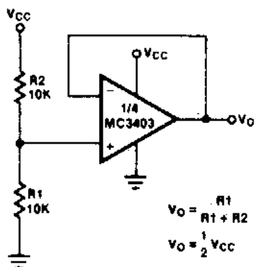


Fig. 93-6

LOW POWER REGULATOR REFERENCE

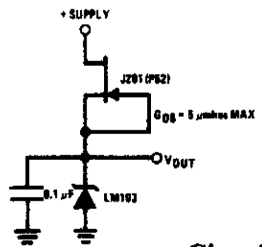


Fig. 93-9

Circuit Notes

This simple reference circuit provides a stable voltage reference almost totally free of supply voltage hash. Typical power supply rejection exceeds 100 dB.

HIGH STABILITY VOLTAGE REFERENCE

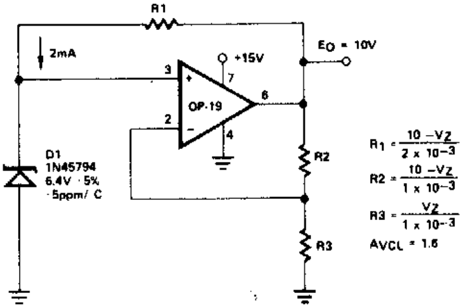


Fig. 93-10

ZENERLESS PRECISION MILLIVOLT SOURCE

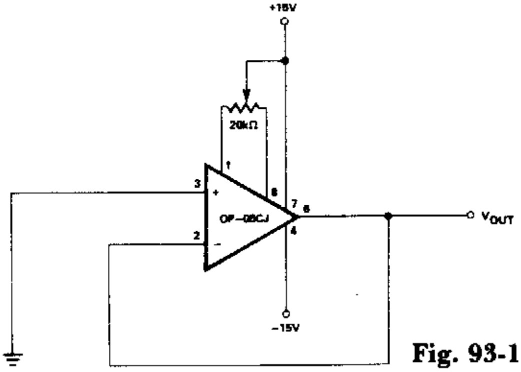


Fig. 93-13

± 3 V REFERENCE

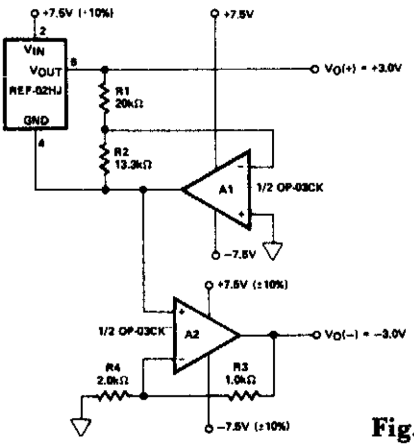


Fig. 93-11

± 10 V REFERENCE

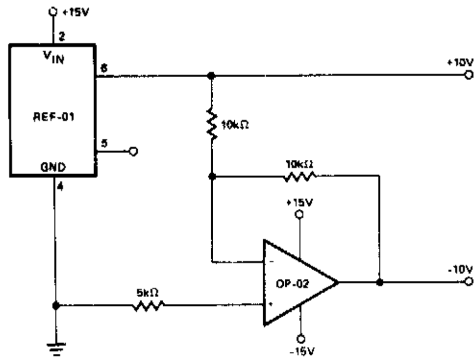


Fig. 93-14

± 5 V REFERENCE

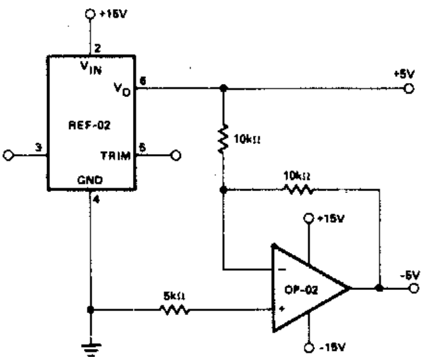


Fig. 93-12

PRECISION REFERENCE SQUARE WAVE VOLTAGE REFERENCE

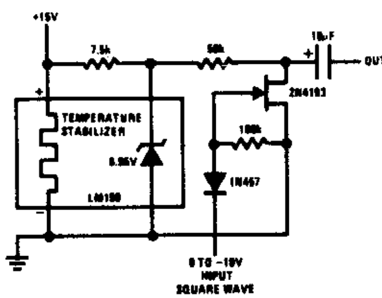
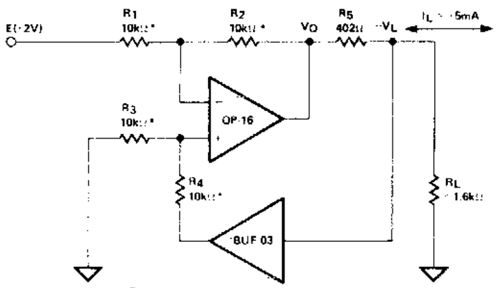


Fig. 93-15

INVERTING BIPOLAR CURRENT SOURCE (HIGH SPEED)



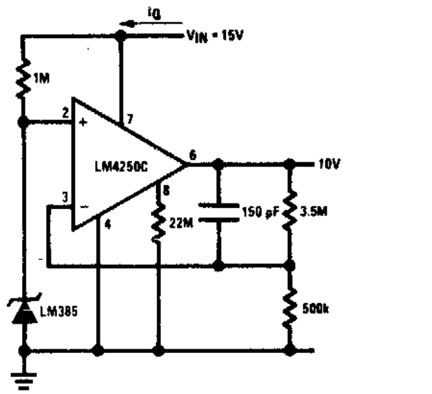
*MATCHED SO THAT $\frac{R_2}{R_1} = \frac{R_4}{R_3}$

$I_L = \frac{E}{R_5} \cdot \frac{R_2}{R_1}$

COMPLIANCE OF ABOVE CIRCUIT (ΔV_L) IS 8V WHEN E = 2V AND R_L = 1.6k. NOTE THAT V_O is 10V UNDER THESE CONDITIONS.

Fig. 93-16

PRECISION REFERENCE MICROPOWER 10 V REFERENCE



* $I_Q \geq 20\ \mu A$ standby current

Fig. 93-17

PRECISION REFERENCE LOW NOISE BUFFERED REFERENCE

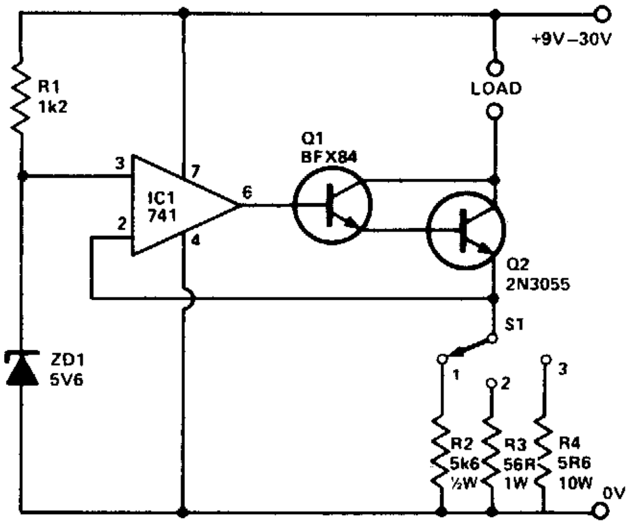


Fig. 93-18

Circuit Notes

The circuit will provide 3 preset currents which will remain constant despite variations of ambient temperature or line voltage. ZD1 produces a temperature stable reference voltage which is applied to the noninverting input of IC1. 100% feedback is applied from the output to the inverting input holding the voltage at Q2s emitter at the same potential as the noninverting input. The current flowing into the load therefore is defined solely by the resistor selected by S1. With the values employed here, a preset current of 10 mA, 100 mA or 1 A can be selected. Q2 should be mounted on a suitable heatsink.

CONSTANT CURRENT SOURCE

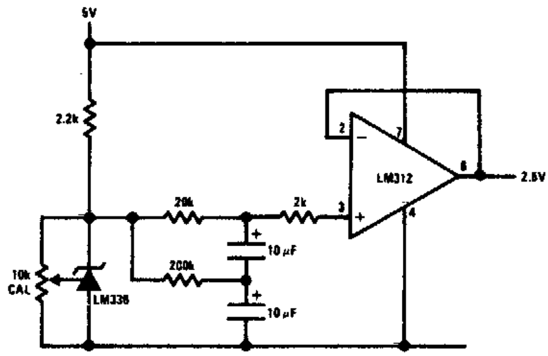


Fig. 93-19

PRECISION DUAL TRACKING VOLTAGE REFERENCES

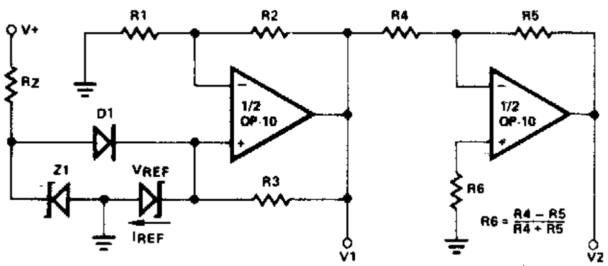


Fig. 93-20

PRECISION REFERENCE BIPOLAR OUTPUT REFERENCE

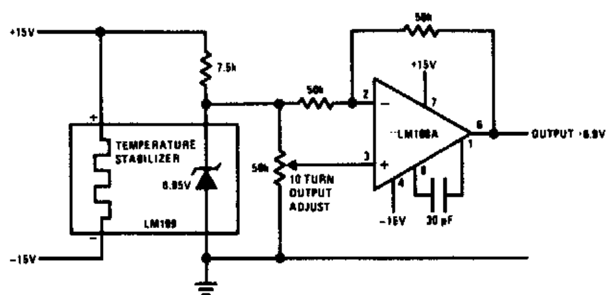


Fig. 93-21

PRECISION REFERENCE 0 V TO 20 V POWER REFERENCE

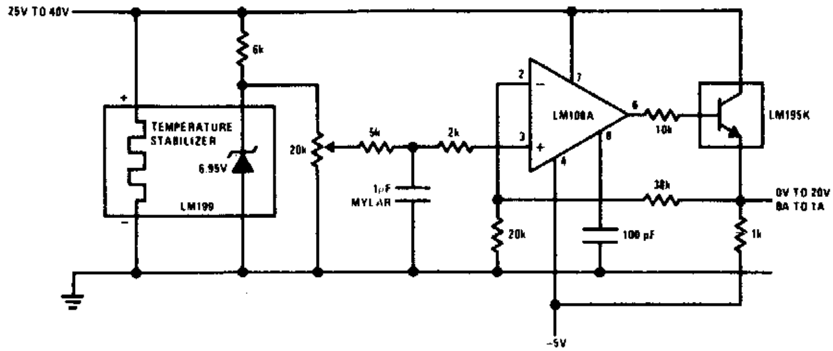


Fig. 93-22

PRECISION REFERENCE STANDARD CELL REPLACEMENT

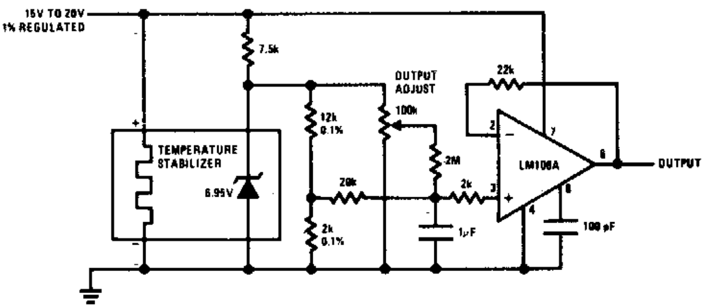


Fig. 93-23

94

Voltage-Controlled Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LINEAR VOLTAGE CONTROLLED OSCILLATOR

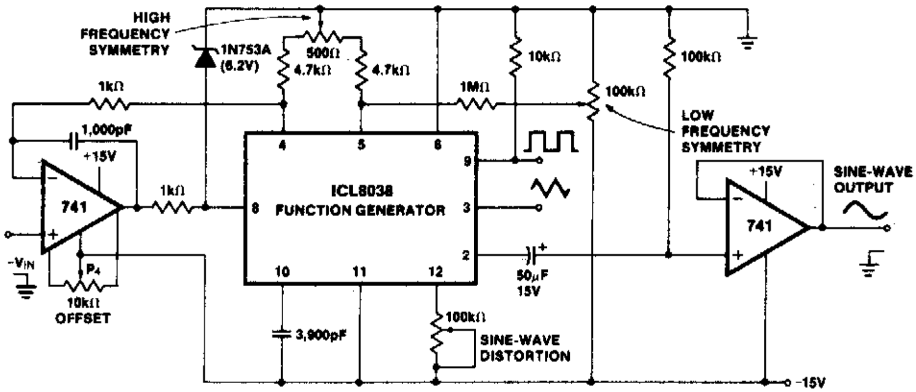


Fig. 94-1

Circuit Notes

The linearity of input sweep voltage versus output frequency is significantly improved by using an op amp.

10 Hz TO 10 kHz VOLTAGE CONTROLLED OSCILLATOR

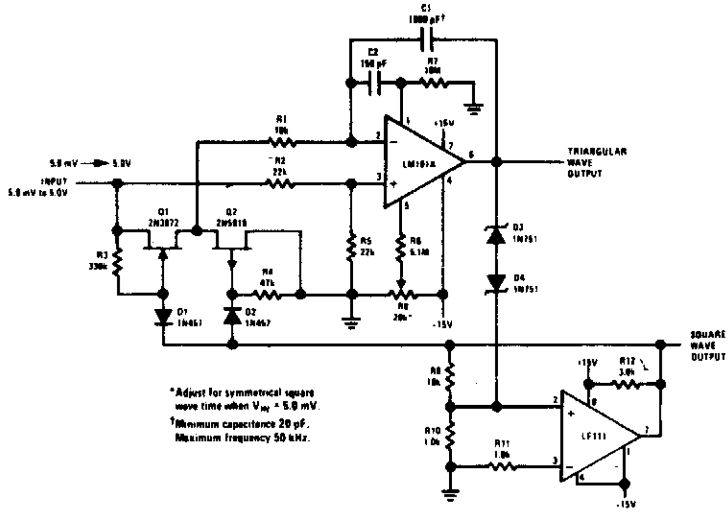


Fig. 94-2

PRECISION VOLTAGE CONTROLLED OSCILLATOR

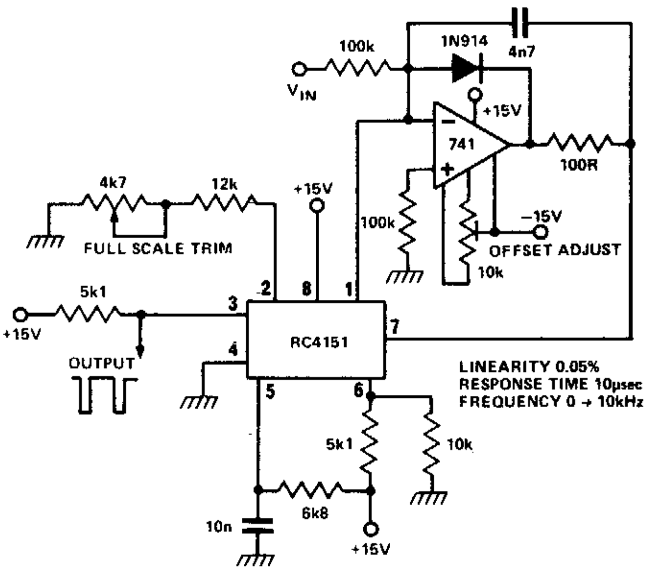
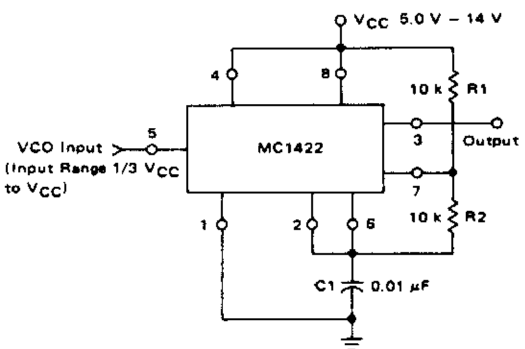


Fig. 94-3

Circuit Notes

RC 4151 precision voltage-to-frequency converter generates a pulse train output linearly proportional to the input voltage.

VOLTAGE CONTROLLED OSCILLATOR



Circuit Notes

The VCO circuit, which has a nonlinear transfer characteristic, will operate satisfactorily up to 200 kHz. The VCO input range is effective from ⅓ Vcc to Vcc − 2 V, with the highest control voltage producing the lowest output frequency.

Fig. 94-4

SIMPLE VOLTAGE CONTROLLED OSCILLATOR

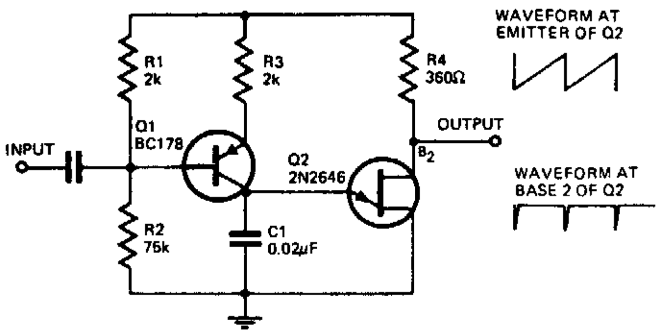


Fig. 94-5

Circuit Notes

With the component values shown, the oscillator has a frequency of 8 kHz. When an input signal is applied to the base of Q1 the current flowing through Q1 is varied, thus varying the time required to charge C1. Due to the phase inversion in Q1 the direction of output frequency change is 180 degrees out of phase with the input signal. The output may be used to trigger a bistable flip-flop.

THREE DECADES VCO

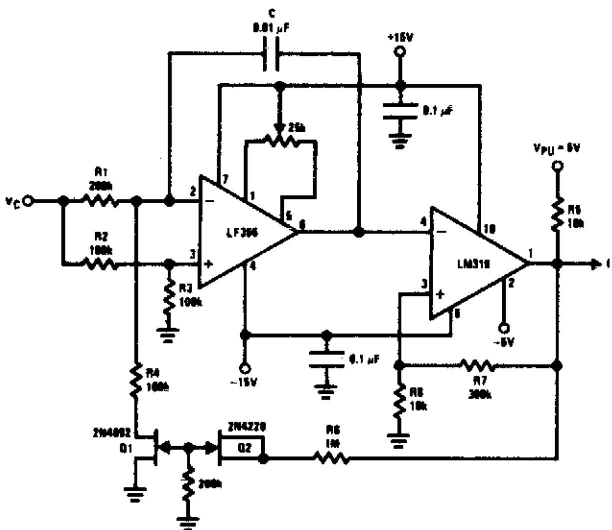
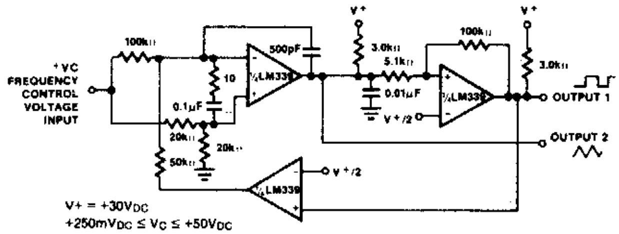


Fig. 94-6

$$f = \frac{V_C\,(R8+R7)}{(8\,V_{PU}\,R8\,R1)\,C},\quad 0 \le V_C \le 30V,\ 10\ Hz \le f \le 10\ kHz$$

R1, R4 matched. Linearity 0.1% over 2 decades.

TWO-DECADE HIGH-FREQUENCY VCO



$V+ = +30V_{DC}$
$+250mV_{DC} \le V_C \le +50V_{DC}$
$700Hz \le f_o \le 100kHz$

Fig. 94-7

VOLTAGE CONTROLLED OSCILLATOR

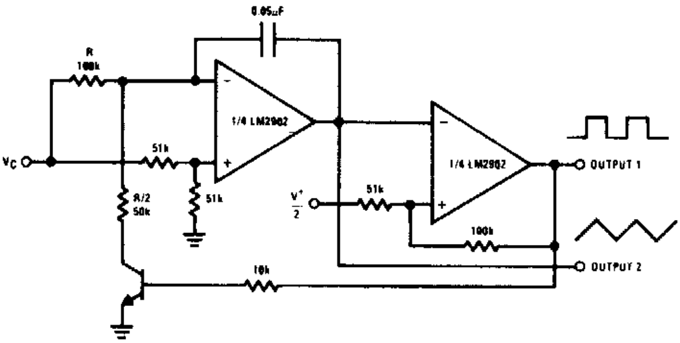
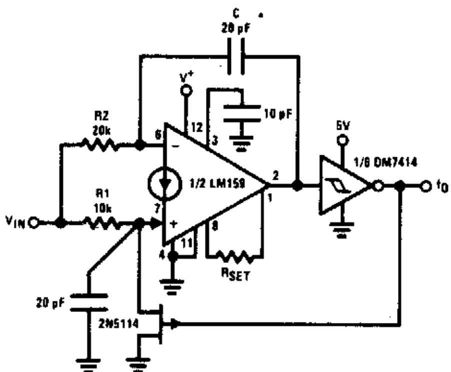


Fig. 94-8

VOLTAGE CONTROLLED OSCILLATOR



$$f_O = \frac{V_{IN} - \phi}{4C\Delta V\ R1}$$

where: $R2 = 2R1$
ϕ = amplifier input voltage ≈ 0.6V
ΔV = DM7414 hysteresis, typ 1V

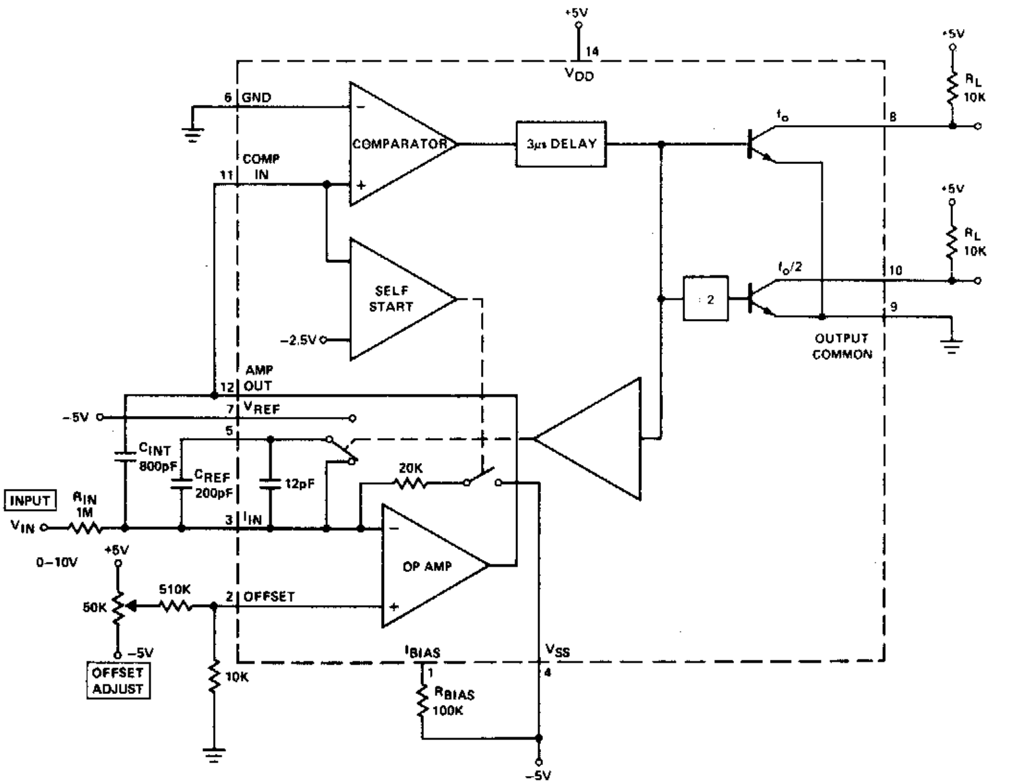
- 5 MHz operation
- T^2L ouput

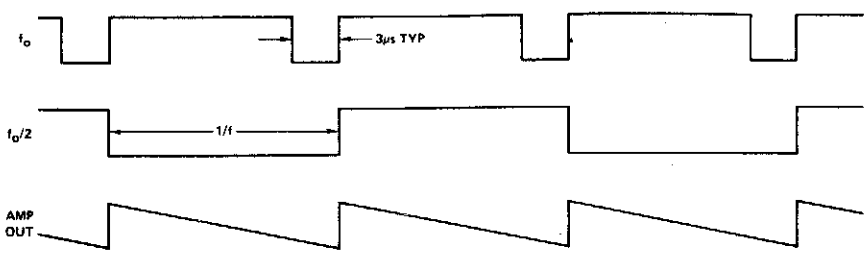
Fig. 94-9

95
Voltage-to-Frequency Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

10 Hz TO 10 kHz VOLTAGE/FREQUENCY CONVERTER





1. To adjust f_{min}, set V_{IN} = 10mV and adjust the 50K offset for 10Hz out.
2. To adjust f_{max}, set V_{IN} 10V and adjust R_{IN} or V_{REF} for 10KHz out.
3. To increase f_{OUT}MAX to 100KHz change C_{REF} to 15pF and C_{INT} to 75pF
4. For high performance applications use high stability components for R_{IN}, C_{REF}, V_{REF} (metal film resistors and glass film capacitors). Also separate the output ground (Pin 9) from the input ground (Pin 6).

Output Waveforms

Fig. 95-1

VOLTAGE-TO-FREQUENCY CONVERTER

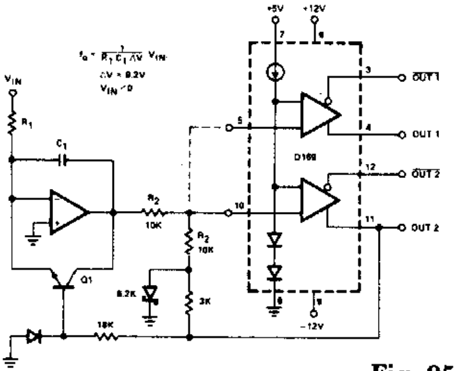


Fig. 95-2

Circuit Notes

The D169 serves as a level detector and provides complementary outputs. The op amp is used to integrate the input signal V_{IN} with a time constant of R1C1. The input (must be negative) causes a positive ramp at the output of the integrator which is summed with a negative zener voltage. When the ramp is positive enough D169 outputs change state and OUT 2 flips from negative to positive. The output pulse repetition rate f_o, is directly proportioned to the negative input voltage V_{IN}.

VOLTAGE-TO-FREQUENCY CONVERTER

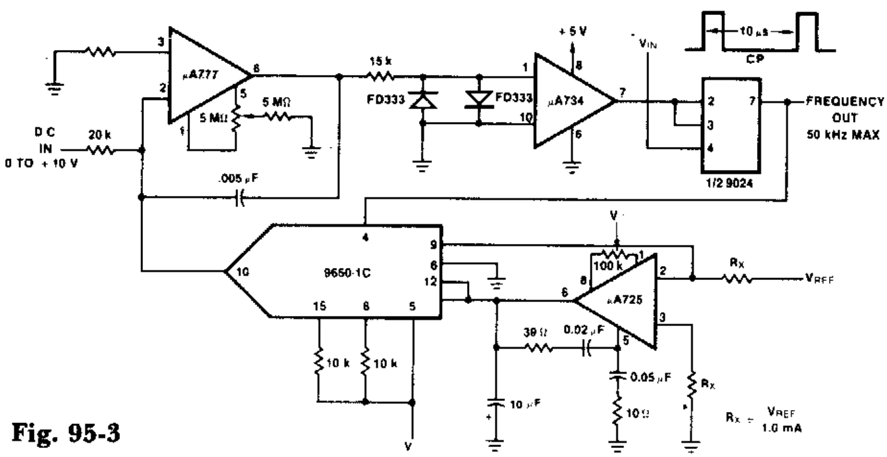


Fig. 95-3

V/F CONVERSION, POSITIVE INPUT VOLTAGE

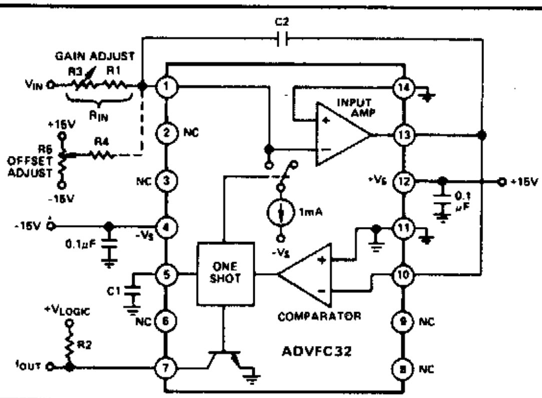
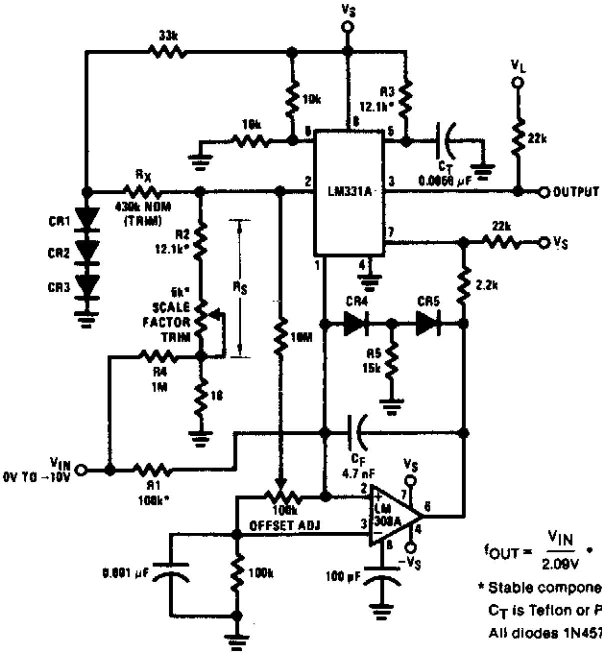


Fig. 95-4

ULTRAPRECISION V/F CONVERTER



$$f_{OUT} = \frac{V_{IN}}{2.09V} \cdot \frac{R_S}{R1} \cdot \frac{1}{R_T C_T}$$

Full-scale output 10 kHz

* Stable components with low tempco; see text

C_T is Teflon or Polystyrene

All diodes 1N457, 1N484, or FD333 (low-leakage silicon)

Fig. 95-5

Circuit Notes

The circuit is capable of better than 0.02% error and 0.003% nonlinearity for a ±20°C range about room temperature.

V/F CONVERSION, NEGATIVE INPUT VOLTAGE

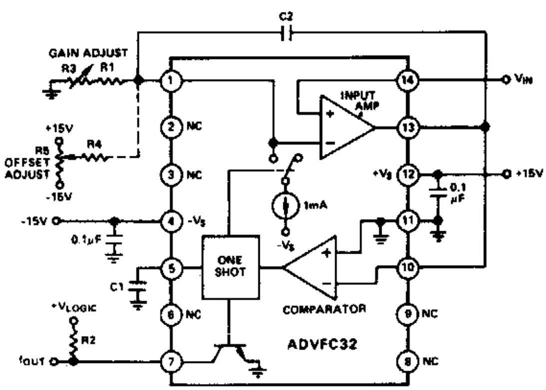
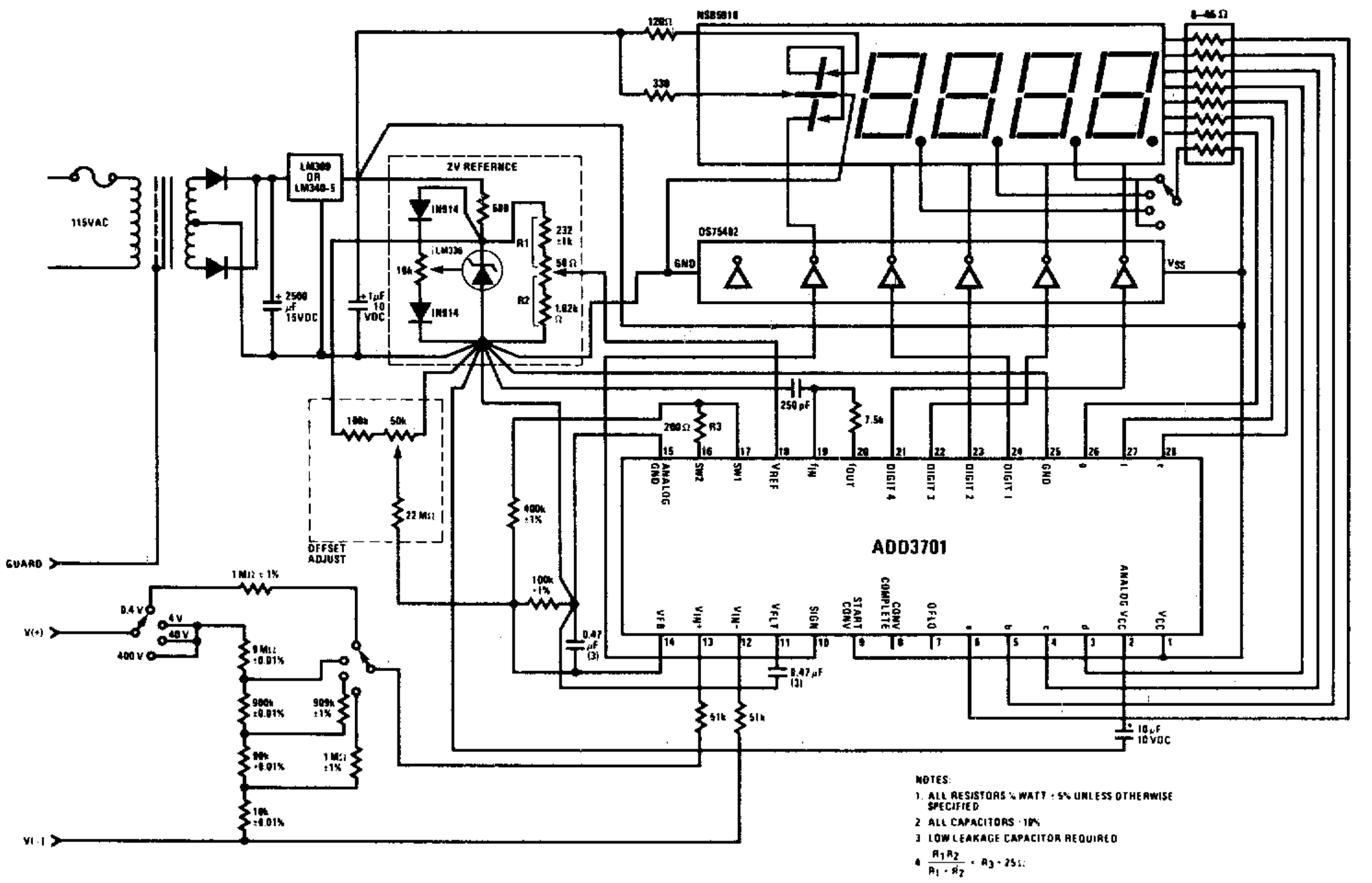


Fig. 95-6

96
Voltmeters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

3-¾ DIGIT DVM, FOUR DECADE, ±0.4 V, ±4 V, ±40 V, AND ±400 V FULL SCALE



NOTES:

1. ALL RESISTORS ¼ WATT ±5% UNLESS OTHERWISE SPECIFIED
2. ALL CAPACITORS ±10%
3. LOW LEAKAGE CAPACITOR REQUIRED
4. $\frac{R_1 R_2}{R_1 + R_2} + R_3 = 25\Omega$

Fig. 96-1

AUTOMATIC NULLING DVM

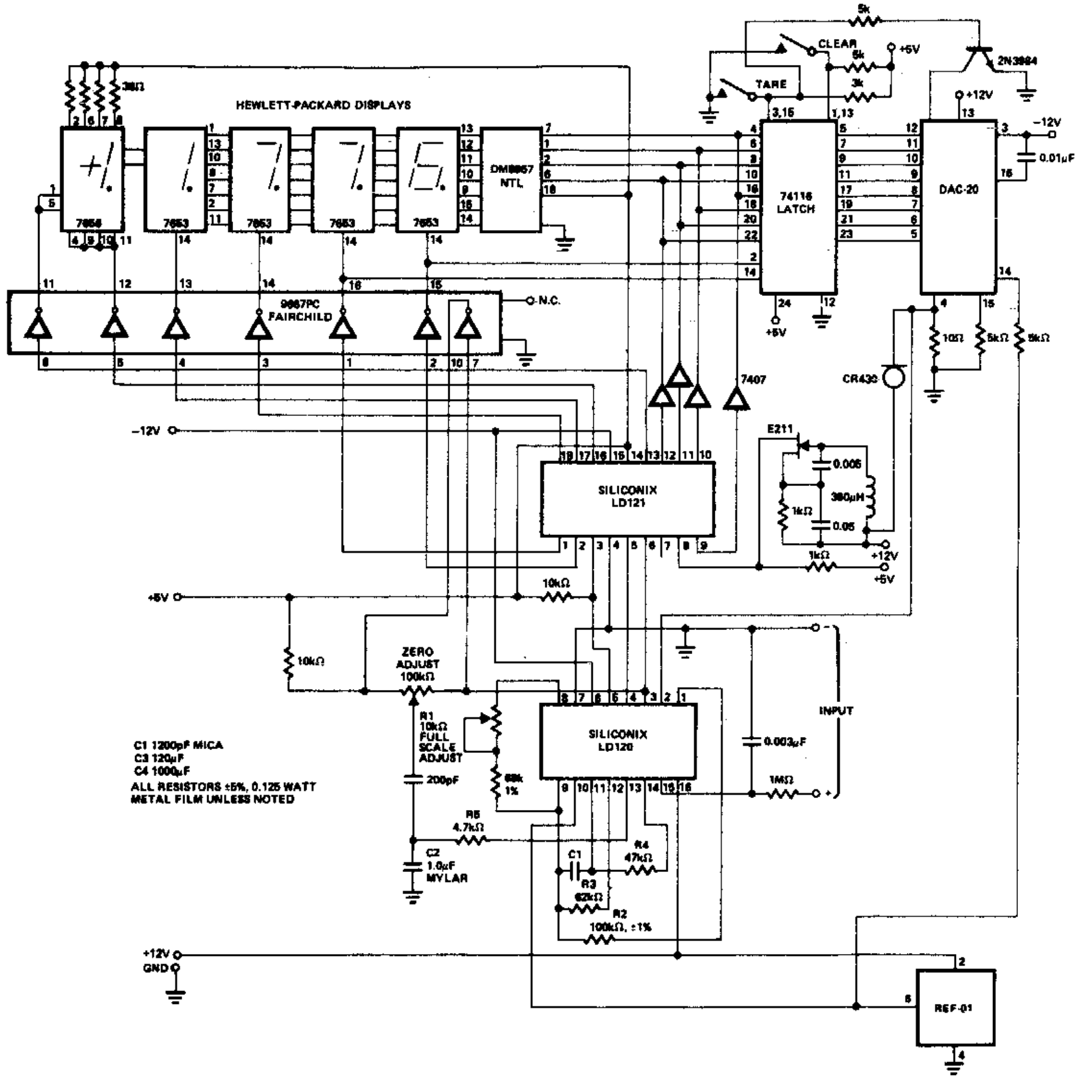


Fig. 96-2

3-½ DIGIT TRUE RMS AC VOLTMETER

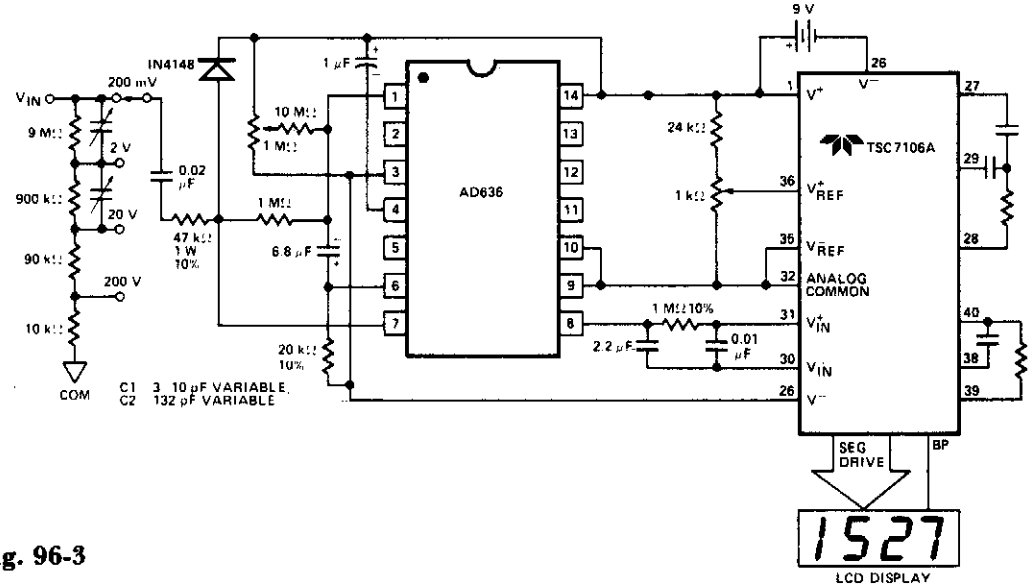


Fig. 96-3

3½ DIGIT DVM (±200.0 mV) COMMON ANODE DISPLAY

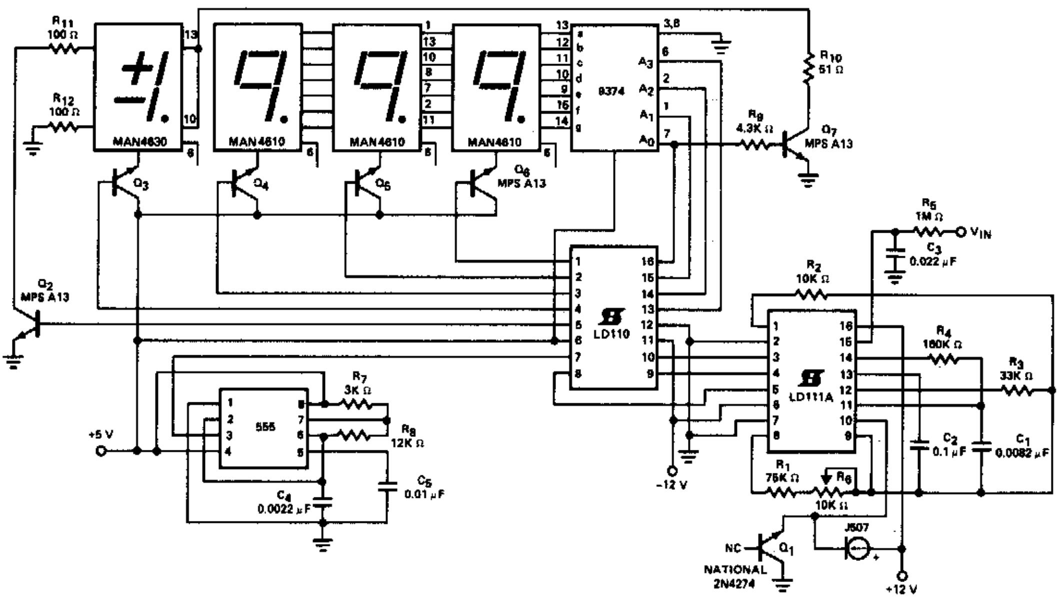


Fig. 96-4

DVM AUTO-CALIBRATE CIRCUIT

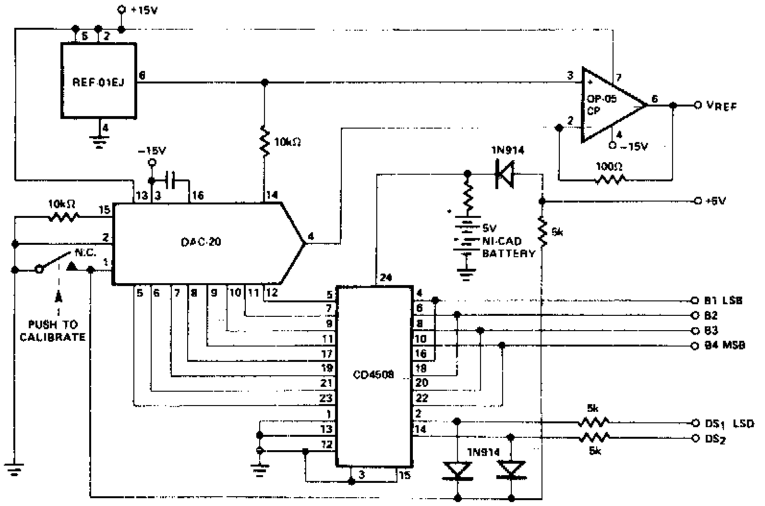


Fig. 96-5

FET VOLTMETER

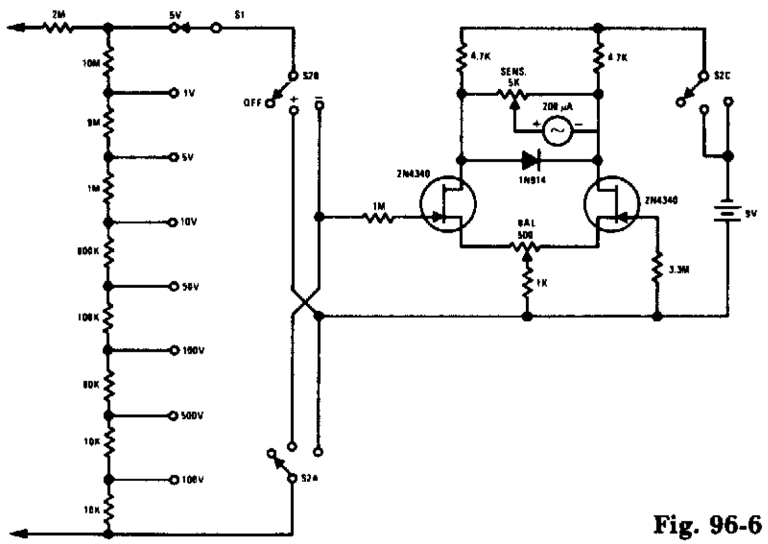
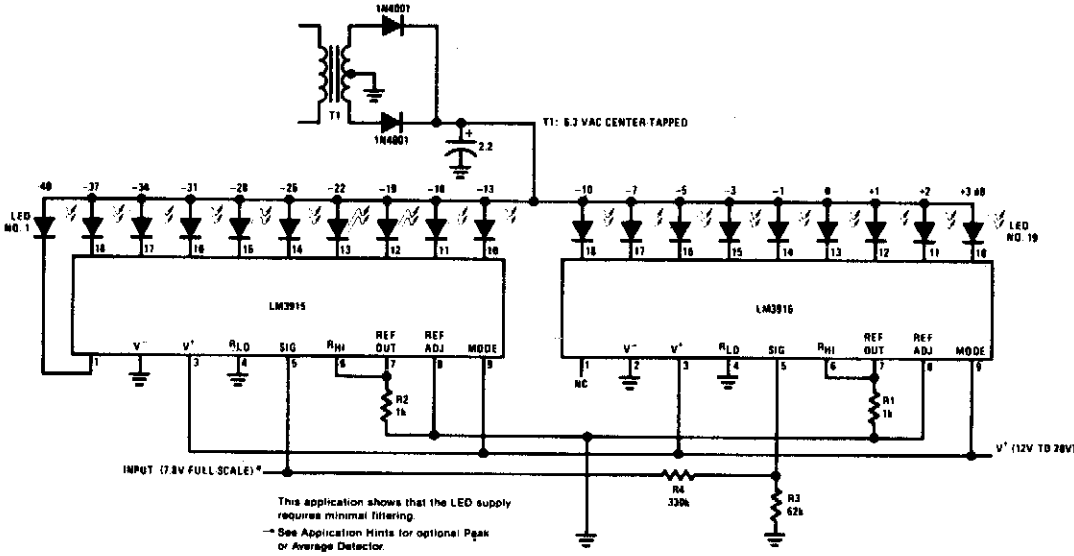


Fig. 96-6

Circuit Notes

This FETVM replaces the function of the VTVM while at the same time ridding the instrument of the usual line cord. In addition, drift rates are far superior to vacuum tube circuits allowing a 0.5 volt full scale range which is impractical with most vacuum tubes. The low-leakage, low-noise 2N4340 is an ideal device for this application.

EXTENDED RANGE VU METER (BAR MODE)



This application shows that the LED supply requires minimal filtering.

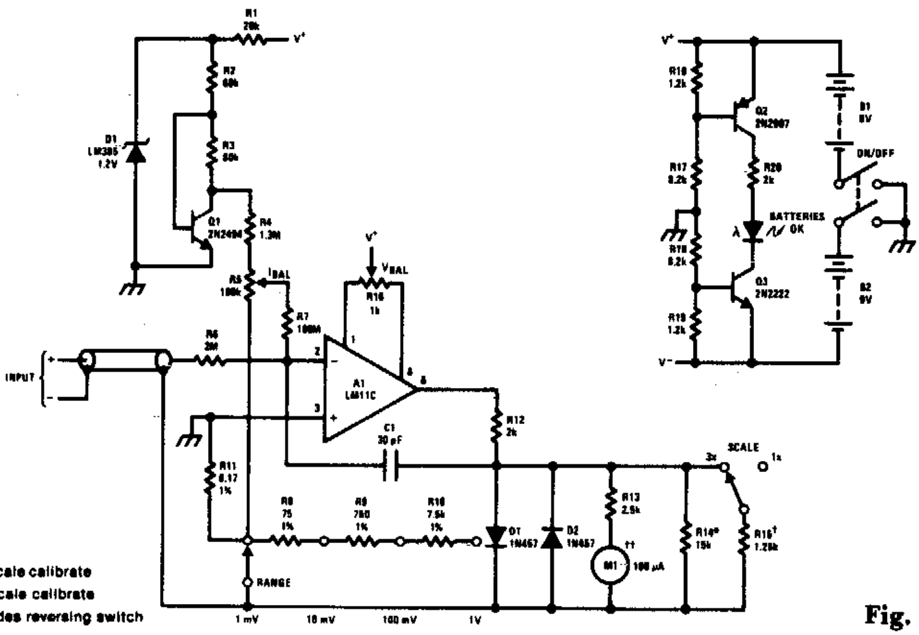
* See Application Hints for optional Peak or Average Detector.

† Adjust R3 for 3 dB difference between LED #11 and LED #12.

$$\frac{R3}{R3 + R4} = 0.158 = -16\text{ dB}$$

Fig. 96-7

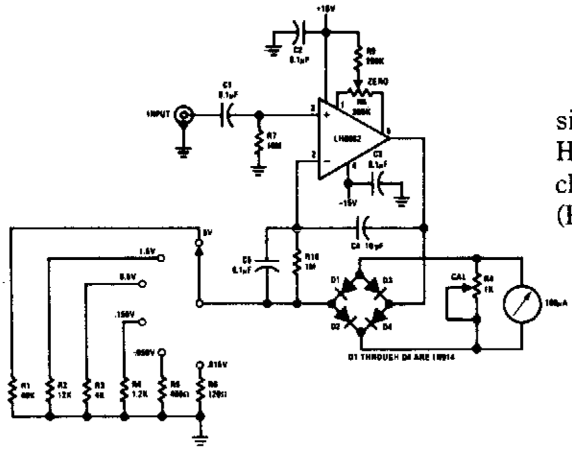
HIGH INPUT IMPEDANCE MILLIVOLTMETER



* 1× scale calibrate
† 3× scale calibrate
‡ includes reversing switch

Fig. 96-8

WIDE BAND AC VOLTMETER



Circuit Notes

This voltmeter is capable of measuring ac signals as low as 15 mV at frequencies from 100 Hz to 500 kHz. Full scale sensitivity may be changed by altering the values R1 through R6 ($R = V_{IN}/100\ \mu A$).

Fig. 96-9

SUPPRESSED ZERO METER

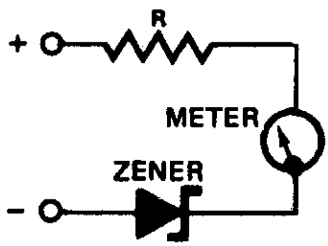


Fig. 96-10

Circuit Notes

A zener diode placed in series with a voltmeter will prevent the meter from reading until the applied voltage exceeds the zener voltage. Thus, a 10 volt zener in series with a 5-volt meter will allow the condition of a 12 V car battery to be monitored with much greater sensitivity than would be possible with a meter reading 0-15 volts.

AC MILLIVOLTMETER

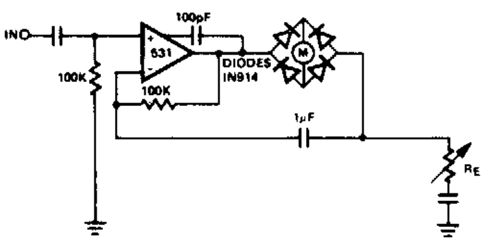


Fig. 96-11

4½ DIGIT LCD-DVM

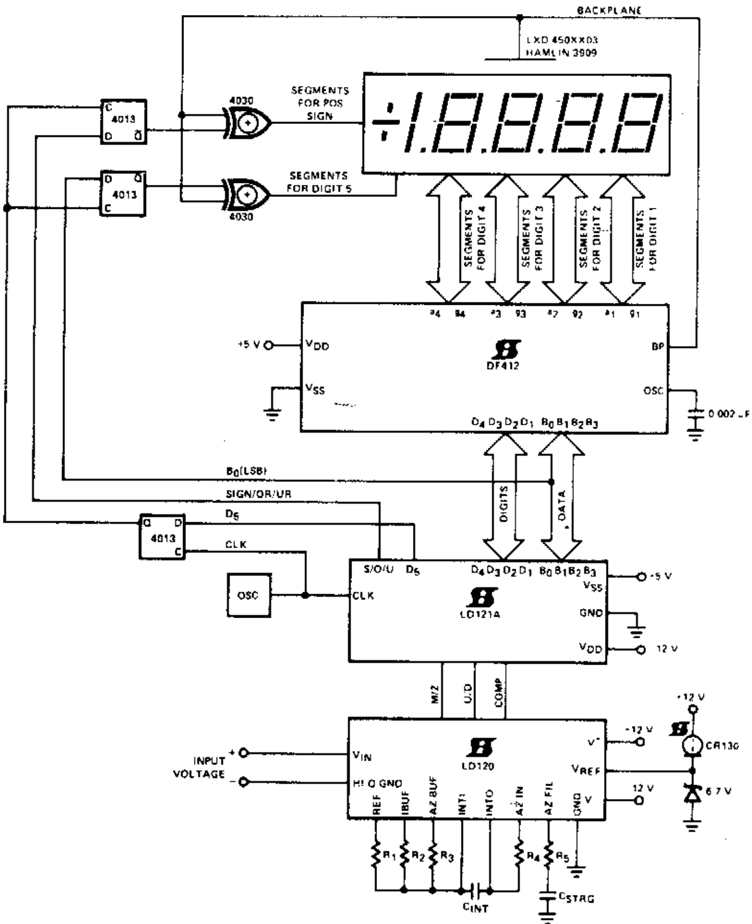


Fig. 96-12

SENSITIVE LOW COST "VTVM"

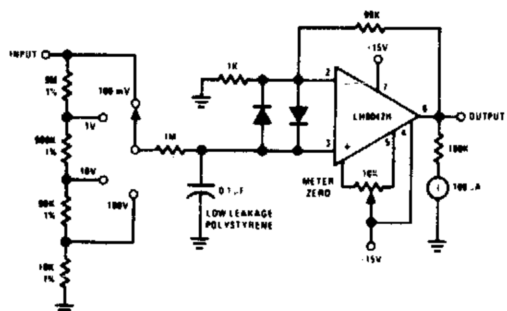


Fig. 96-13

97

Waveform and Function Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW COST ADJUSTABLE FUNCTION GENERATOR

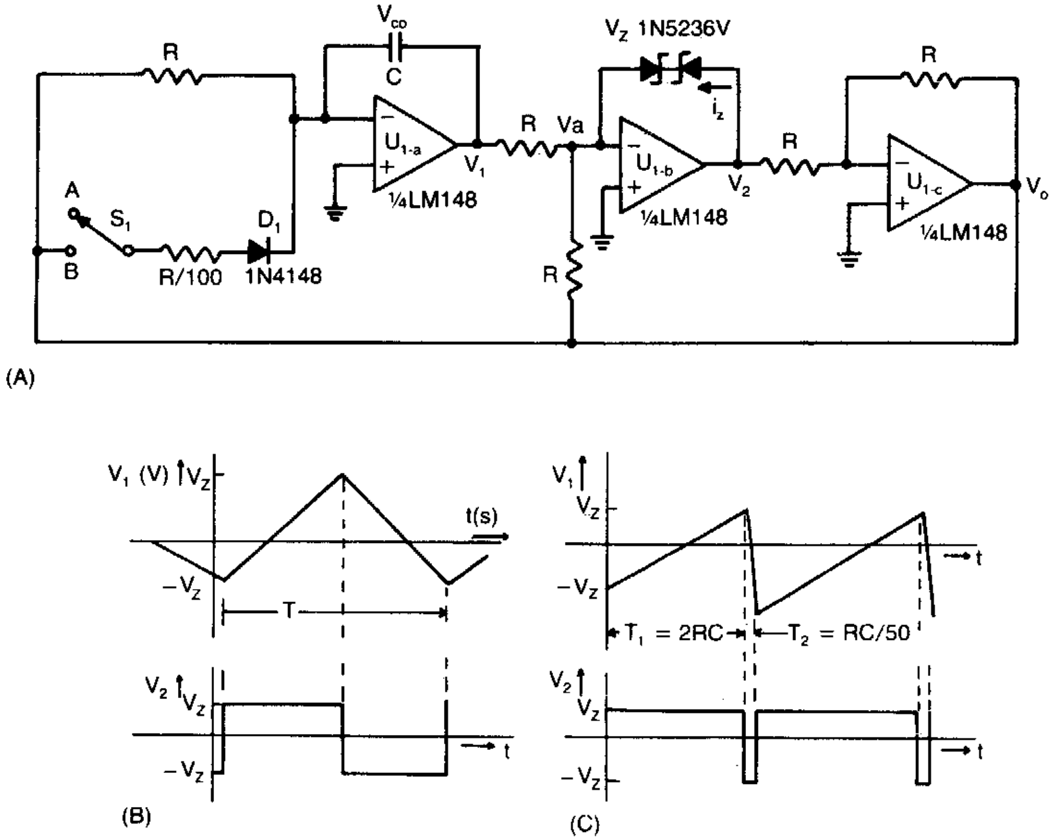
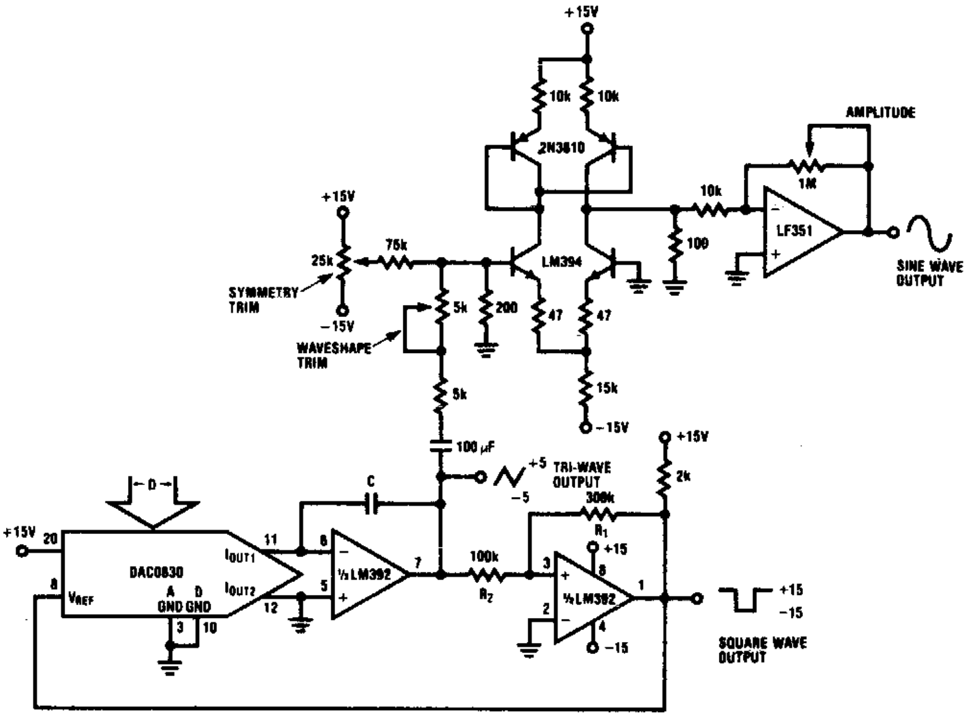


Fig. 97-1

Circuit Notes

This low-cost operational-amplifier circuit (A) generates four different functions with adjustable periods. For the components shown here, the period of the output waveforms is given by T = 4RC and T = 2RC. With switch S1 in position A, V1 is a triangular waveform, while V2 is a square wave (B). With the switch in position B, a sawtooth waveform is generated at V1 and a pulse at V2(C).

DAC CONTROLLED FUNCTION GENERATOR



- DAC controls the frequency of sine, square, and triangle outputs.
- $f = \frac{D}{256(20k)C}$ for $V_{OMAX} = V_{OMIN}$ of square wave output and $R_1 = 3R_2$.
- 255 to 1 linear frequency range; oscillator stops with D = 0
- Trim symmetry and wave-shape for minimum sine wave distortion.

Fig. 97-2

PROGRAMMED FUNCTION GENERATOR

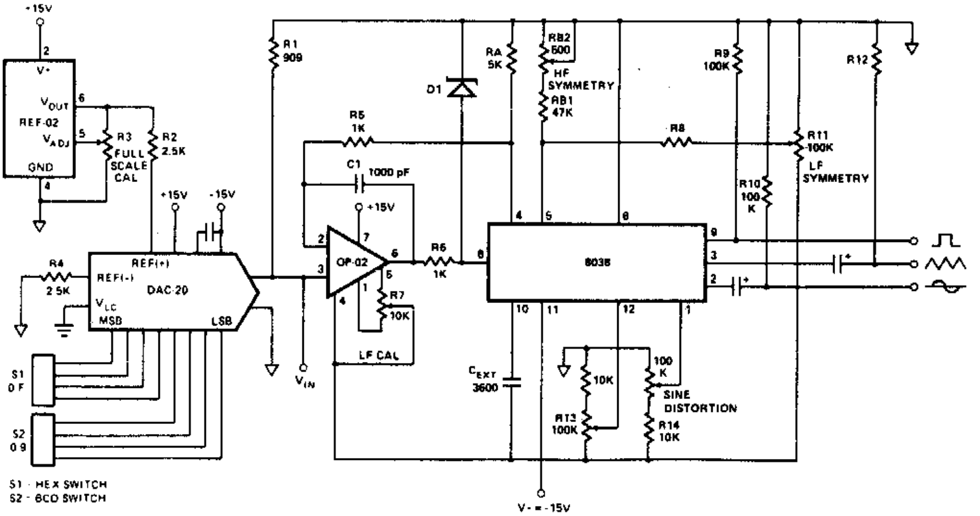
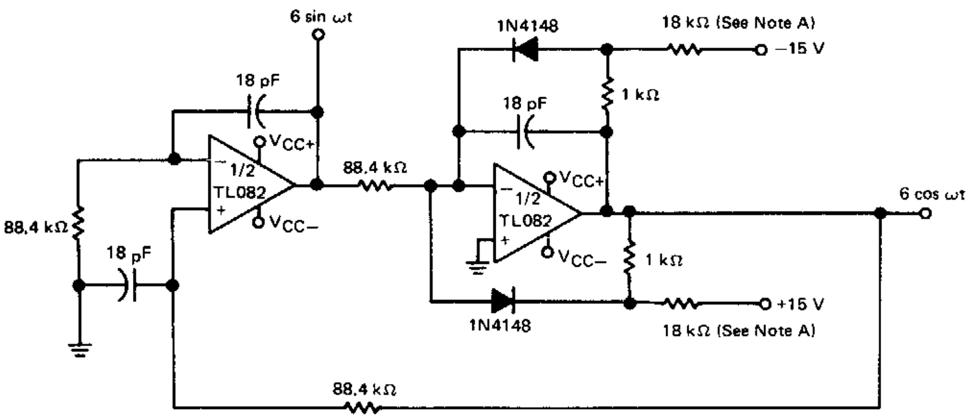


Fig. 97-3

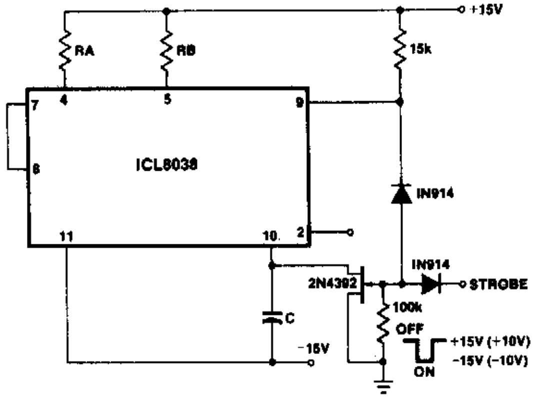
100-kHz QUADRATURE OSCILLATOR



Note A: These resistor values may be adjusted for a symmetrical output.

Fig. 97-4

STROBE-TONE BURST GENERATOR

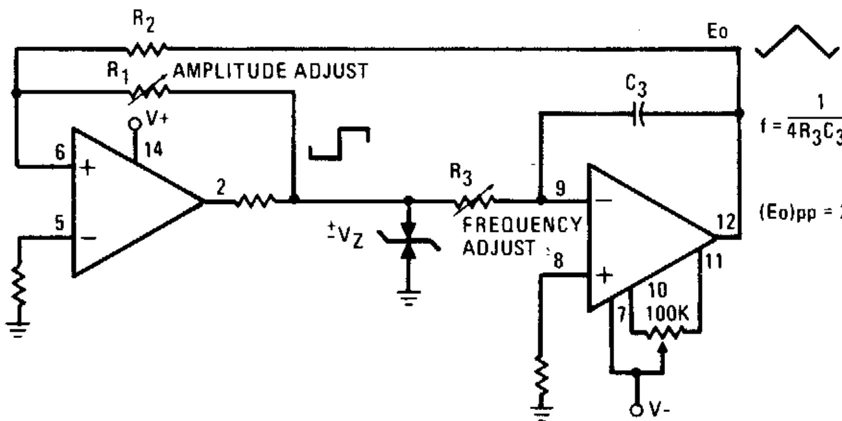


Circuit Notes

With a dual supply voltage, the external capacitor on pin 10 can be shorted to ground to halt the 8038 oscillation. The circuit uses a FET switch and diode ANDed with an input strobe signal to allow the output to always start on the same slope.

Fig. 97-5

LOW COST HIGH FREQUENCY GENERATOR



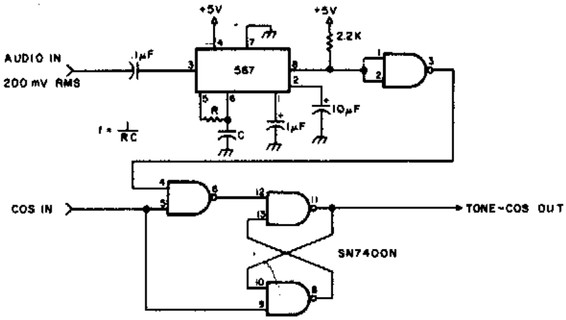
$$f = \frac{1}{4R_3C_3}\left(\frac{R_1}{R_2}\right)$$

$$(E_o)_{pp} = 2V_Z\left(\frac{R_2}{R_1}\right)$$

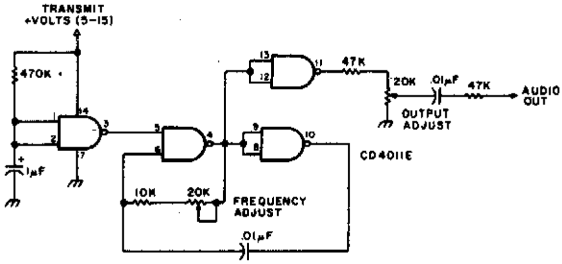
IC =HA2650/26555

Fig. 97-6

TONE-BURST OSCILLATOR AND DECODER



Decoder and logic.



Tone-burst oscillator.

Fig. 97-7

Circuit Notes

A tone burst sent at the beginning of each transmission is decoded (at receiver) by a PLL causing output from pin 3 of logic gate to turn on carrier-operated switch (COS).

TRIANGLE AND SQUARE WAVEFORM GENERATOR

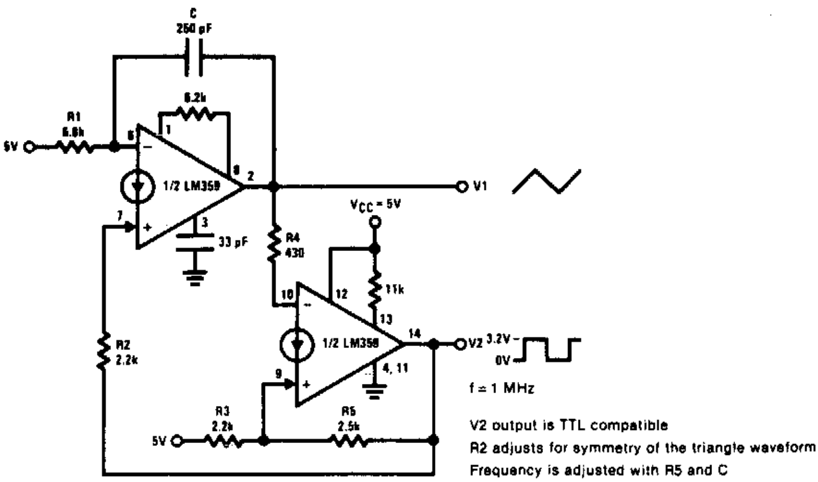
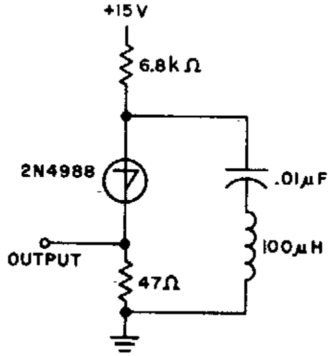


Fig. 97-8

10 kHz OSCILLATOR



Circuit Notes

The capacitor charges until switching voltage is reached. When SUS switches on, the inductor causes current to ring. When the current thru SUS drops below the holding current, the device turns off and the cycle repeats.

Fig. 97-9

50 kHz OSCILLATOR

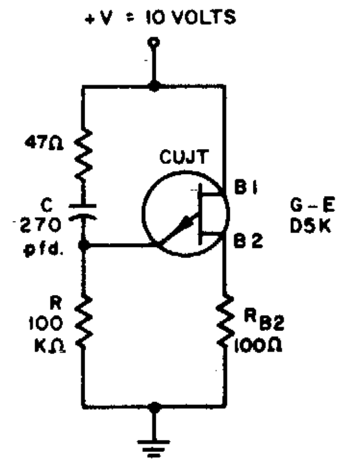


Fig. 97-10

Circuit Notes

A 50 kHz circuit is possible because of the more nearly ideal characteristics of the D5K.

VARIABLE AUDIO OSCILLATOR, 20 Hz TO 20 kHz

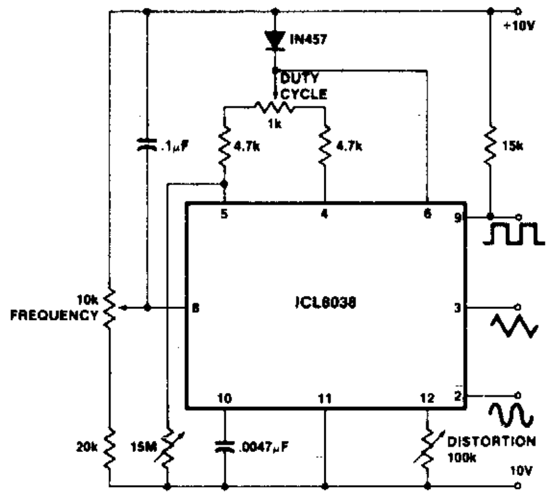


Fig. 97-11

Circuit Notes

To obtain a 1000:1 Sweep Range, the voltage across external resistors R_A and R_B must decrease to nearly zero. This requires that the highest voltage on control pin 8 exceed the voltage at the top of R_A and R_B by a few hundred millivolts. The circuit achieves this by using a diode to lower the effective supply voltage on the 8038. The large resistor on pin 5 helps reduce duty cycle variations with sweep.

GATED OSCILLATOR

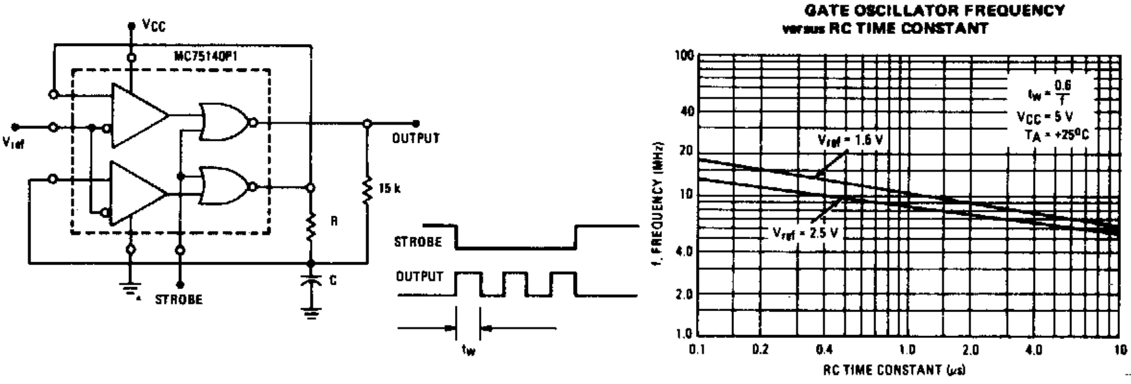


Fig. 97-12

EXPONENTIAL DIGITALLY-CONTROLLED OSCILLATOR

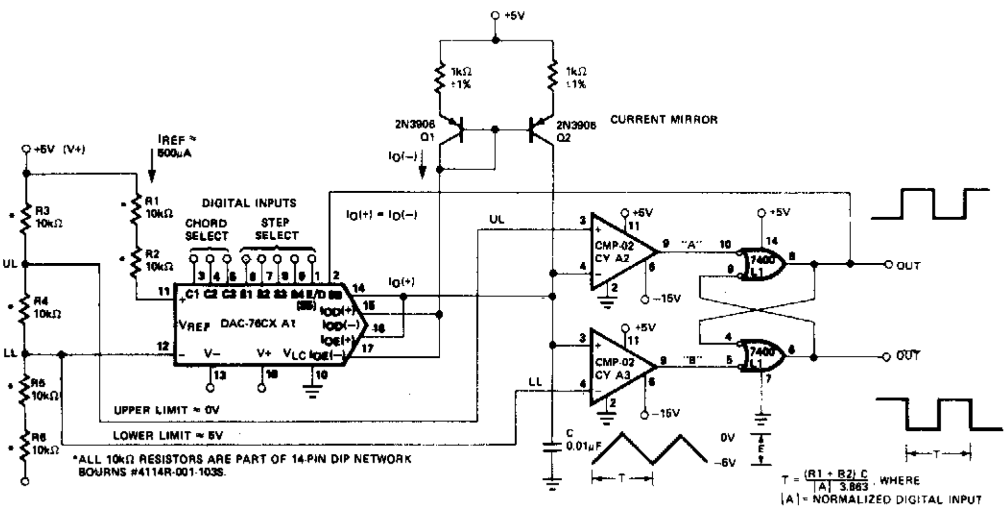


Fig. 97-13

Circuit Notes

The microprocessor-controlled oscillator has a 8159 to 1 frequency range covering 2.5 Hz to 20 kHz. An exponential, current output IC DAC functioning as a programmable current source alternately charges and discharges a capacitor between precisely-controlled upper and lower limits. The circuit features instantaneous frequency change, operates with +5 ±1 V and −15 V ±3 V supplies, and has the dynamic range of a 13-bit DAC.

FUNCTION GENERATOR

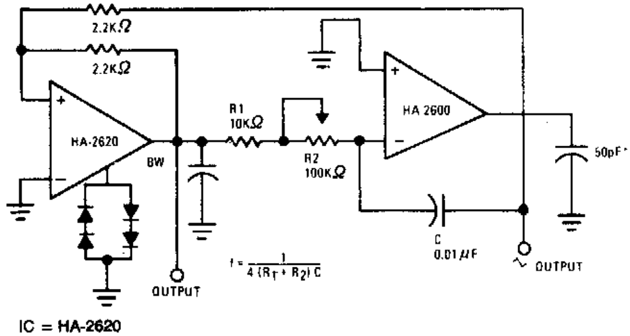
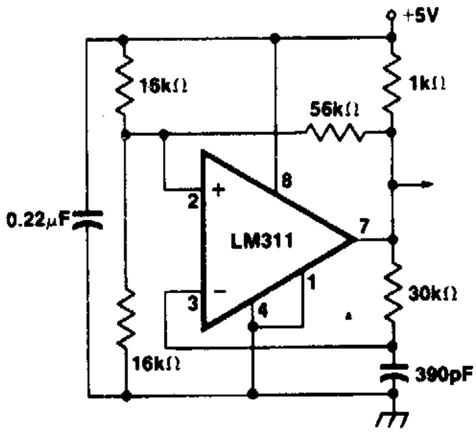


Fig. 97-14

CLOCK SOURCE



Circuit Notes

A clock source using LM311 voltage comparator in positive feedback mode to minimize clock frequency shift problem.

Fig. 97-15

PRECISION OSCILLATOR WITH 20 NS SWITCHING

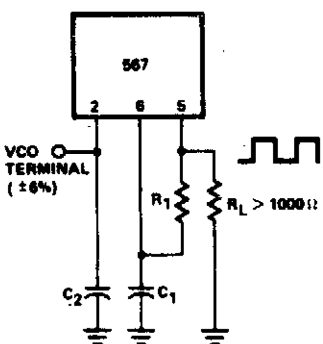


Fig. 97-16

OSCILLATOR WITH QUADRATURE OUTPUT

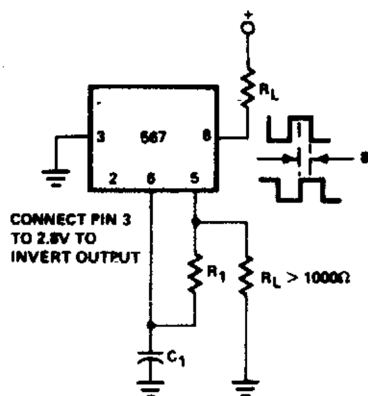


Fig. 97-17

WIDE RANGE VARIABLE OSCILLATOR

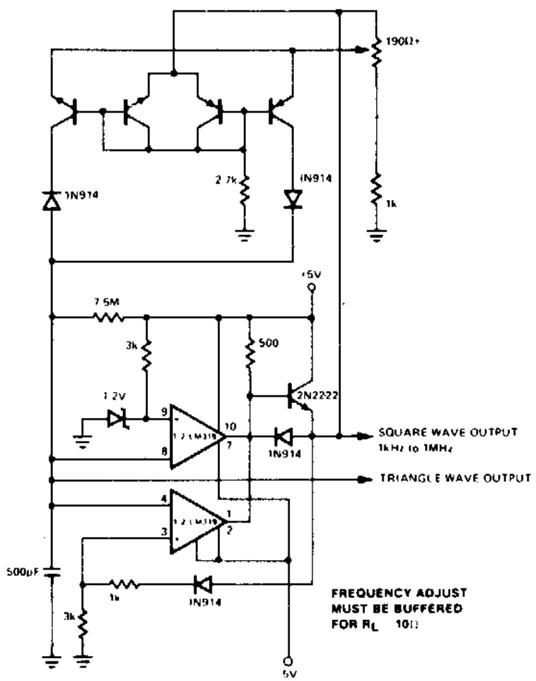
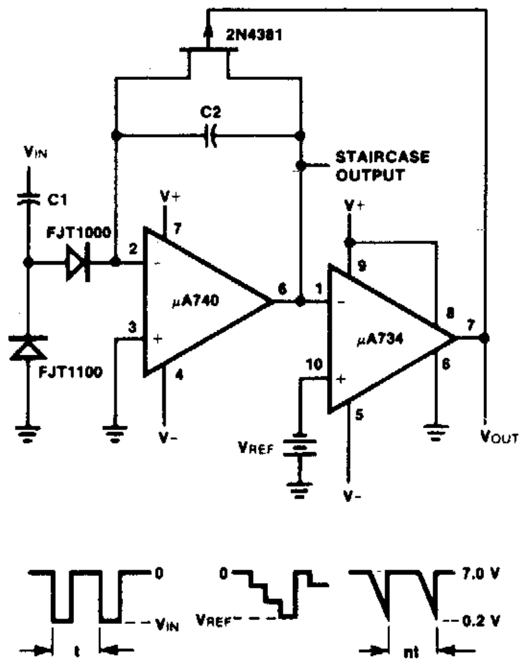


Fig. 97-18

FREQUENCY DIVIDER AND STAIRCASE GENERATOR



$$|V_{REF}| = 2V_D + N\left[3.5T + 2V_D - \frac{C_1 V_{IN}}{C_2}\right]$$

T in Seconds

V_D for FJT 1000 ≈ 0.31 V

Fig. 97-19

PRECISION OSCILLATOR TO SWITCH 100 mA LOADS

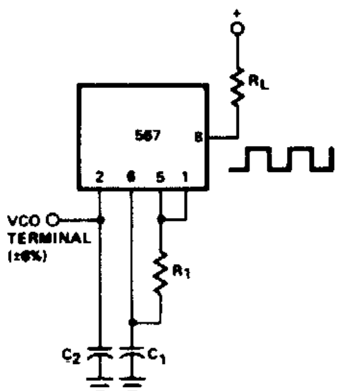


Fig. 97-20

98

Zero Crossing Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Zero Crossing Switch
Zero Crossing Detector
Zero Crossing Detector
Zero Crossing Detector with Temperature Sensor
Zero Crossing Detector
Zero Crossing Detector

ZERO CROSSING SWITCH

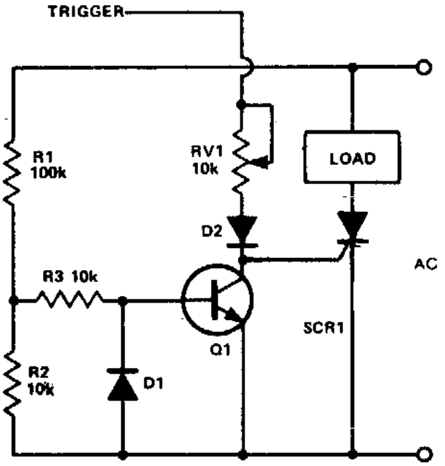


Fig. 98-1

Circuit Notes

When switching loads with the aid of a thyristor, a large amount of RFI can be generated unless some form of zero crossing switch is used. The circuit shows a simple single transistor zero crossing switch. R1 and R2 act as a potential divider. The potential at their junction is about 10% of the ac voltage. This voltage level is fed, via R3, to the transistor's base. If the voltage at this point is above 0.2, the transistor will conduct, shunting any thyristor gate current to ground. When the line potential is less than about 2 V, it is possible to trigger the thyristor. The diode D1 is to remove any negative potential that might cause reverse breakdown.

ZERO CROSSING DETECTOR

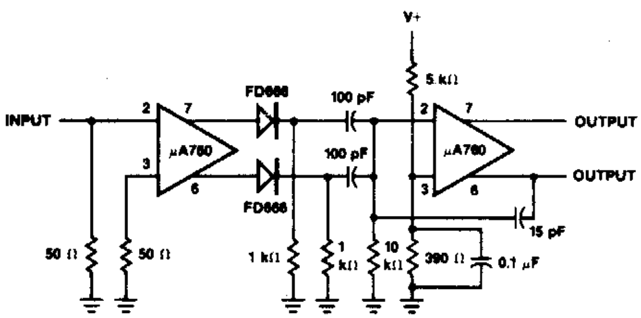
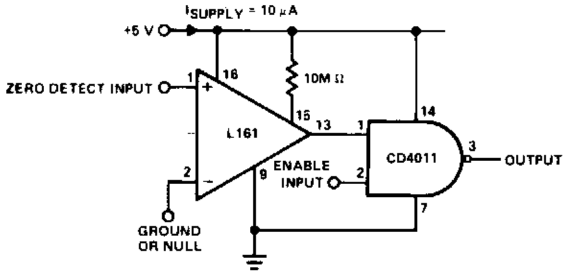


Fig. 98-2

Total Delay = 30 ns
Input frequency = 300 Hz to 3 MHz
Minimum input voltage = 20 mVpk-pk

ZERO CROSSING DETECTOR

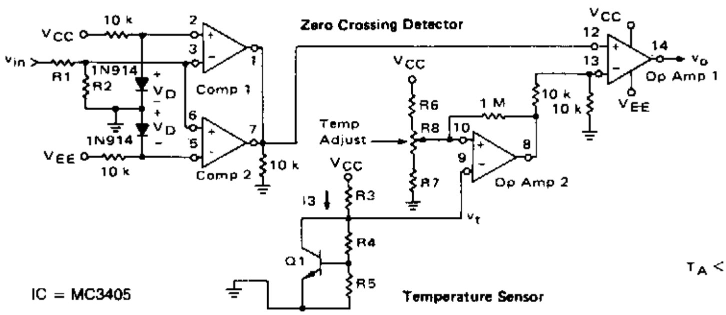


Circuit Notes

This detector is useful in sine wave squaring circuits and A/D converters. The positive input may either be grounded or connected to a nulling voltage which cancels input offsets and enables accuracy to within microvolts of ground. The CMOS output will switch to within a few millivolts of either rail for an input voltage change of less than 200 μV.

Fig. 98-3

ZERO CROSSING DETECTOR WITH TEMPERATURE SENSOR



$v_t = (V_{BE}\text{ of Q1})\left(\frac{R4 + R5}{R5}\right)$

$I3 > \frac{2V_{BE}}{R5}$

R1 and R2 control the switching voltage of the zero crossing detector

$\pm V_s = \pm V_D\,\frac{R1 + R2}{R2}$

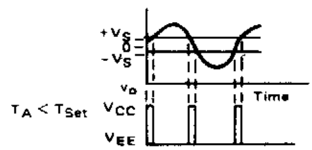


Fig. 98-4

ZERO-CROSSING DETECTOR

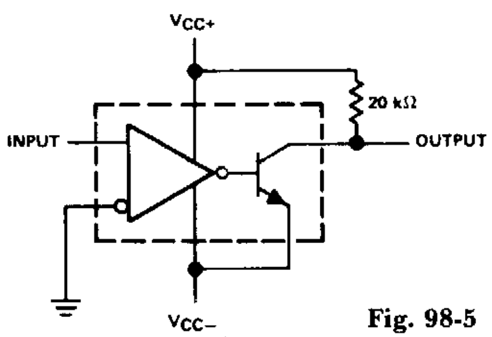


Fig. 98-5

ZERO CROSSING DETECTOR

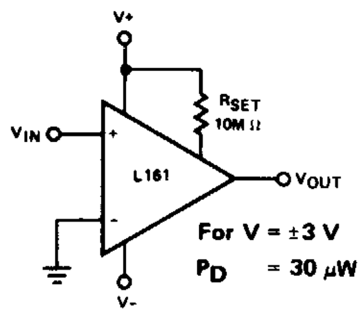


Fig. 98-6

Sources

Chapter 1

Fig. 1-1: The Build-It Book Of Electronics Projects, TAB Book No. 1498, p. 73.
Fig. 1-2: QST, 7/81, p. 28.
Fig. 1-3: Radio Electronics, 10/78, p. 41.
Fig. 1-4: '73 Magazine, 10/77, p. 122.
Fig. 1-5: Modern Electronics, 2/78, p. 50.
Fig. 1-6: Electronics Today International, 3/82, p. 69.
Fig. 1-7: Modern Electronics, 7/78, p. 51.
Fig. 1-8: Electronics Today International, 4/83, p. 72.
Fig. 1-9: 101 Electronic Projects, 1977, #64.
Fig. 1-10: Electronics Today International, 10/78, p. 94.
Fig. 1-11: Modern Electronics, 2/78, p. 55.
Fig. 1-12: Modern Electronics, 2/78, p. 48.
Fig. 1-13: Signetics 555 Timers, 1973, p. 26.
Fig. 1-14: Electronics Today International, 3/83, p. 23.
Fig. 1-15: Electronics Today International, 3/83, p. 23.
Fig. 1-16: National Semiconductor, Linear Databook, 1982, p. 3-288.
Fig. 1-17: Electronics Today International, 3/83, p. 23.
Fig. 1-18: Signetics 555 Timers, 1973, p. 22.
Fig. 1-19: 101 Electronic Projects, 1977, #65.
Fig. 1-20: Modern Electronics, 6/78, p. 58.
Fig. 1-21: Modern Electronics, 6/78, p. 55.

Chapter 2

Fig. 2-1: Modern Electronics, 3/78, p. 69.
Fig. 2-2: Electronics Today International, 10/78, p., 30.
Fig. 2-3: CQ, 5/77, p. 50.
Fig. 2-4: Ham Radio, 10/78, p. 34.
Fig. 2-5: Ham Radio, 10/78, p. 89.
Fig. 2-6: 73 Magazine, 7/78, p. 62.
Fig. 2-7: 101 Electronic Projects, 1975, p. 22.
Fig. 2-8: 73 Magazine, 7/82, p. 46.
Fig. 2-9: 73 Magazine, 7/83, p. 103.
Fig. 2-10: 101 Electronic Projects, 1975, p. 13.
Fig. 2-11: Ham Radio, 5/78, p. 87.
Fig. 2-12: 73 Magazine, p. 164.
Fig. 2-13: Modern Electronics, 2/78, p. 16.
Fig. 2-14: 73 Magazine, 10/77, p. 52.
Fig. 2-15: 73 Magazine, 7/77, p. 34.
Fig. 2-16: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 120.
Fig. 2-17: Ham Radio, 10/70, p. 76.
Fig. 2-18: Electronics Today International, 7/77, p. 72.

Chapter 3

Fig. 3-1: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-119.
Fig. 3-2: Signetics Analog Data Manual, 1982, p. 3-83.
Fig. 3-3: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-207.
Fig. 3-4. Signetics Analog Data Manual, 1983, p. 10-99.
Fig. 3-5: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-107.
Fig. 3-6: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.
Fig. 3-7: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-67.
Fig. 3-8: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 7-7.
Fig. 3-9: Electronics Today International, 2/82, p. 58.
Fig. 3-10. Signetics Analog Data Manual, 1983, p. 10-100.
Fig. 3-11: Precision Monolithics Incorporated 1981 Full Line Catalog, p. 12-50.
Fig. 3-12: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 9-17.
Fig. 3-13: Signetics Analog Data Manual, 1977, p. 35.
Fig. 3-14: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-39.
Fig. 3-15: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-10.
Fig. 3-16: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B, p. 8-21.
Fig. 3-17: Signetics Analog Data Manual, 1983, p. 17-17.
Fig. 3-18: Intersil Data Book, 5/83, p. 5-36.
Fig. 3-19: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-17.
Fig. 3-20: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-83.
Fig. 3-21: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-160.
Fig. 3-22: Signetics Analog Data Manual, 1982, p. 3-103.
Fig. 3-23: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-127.
Fig. 3-24: Courtesy of Motorola Inc., Linear Integrated Circuits, 1979, p. 3-83.
Fig. 3-25: Courtesy of Motorola Inc.

Linear Integrated Circuits, 1979, p. 3-131.
Fig. 3-26: Harris Semiconductor, Analog Data Book 1984.
Fig. 3-27: Intersil Data Book, 5/83, p. 5-36.
Fig. 3-28: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-37.
Fig. 3-29: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-31.
Fig. 3-30: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-21.
Fig. 3-31: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-15.
Fig. 3-32: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-37.
Fig. 3-33: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 7-56.
Fig. 3-34: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-119.
Fig. 3-35: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-42.
Fig. 3-36: Courtesy of Motorola Inc., Linear Integrated Circuits, p. 3-17.
Fig. 3-37: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-23.
Fig. 3-38: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 145.
Fig. 3-39: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-83.
Fig. 3-40: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-41.
Fig. 3-41: Canadian Projects Number 1, Spring/78, p. 29.
Fig. 3-42: Reprinted with the permission of National Semiconductor Corp. Application Note AN125, p. 2.
Fig. 3-43: Harris Semiconductor, Linear & Data Acquisition Products, p. 2-58.
Fig. 3-44: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook 1982, p. 4-98.
Fig. 3-45: Reprinted with the permission of National Semiconductor Corp. Application Note AN125, p. 3.

Chapter 4

Fig. 4-1: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-8.
Fig. 4-2: Intersil Data Book, 5/83, p. 4-83.
Fig. 4-3: Ferranti, Technical Handbook Vol. 10, Data Converters, 1983, p. 7-10.
Fig. 4-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-12.
Fig. 4-5: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-241.
Fig. 4-6: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-13.
Fig. 4-7: Reprinted with the permission of National Semiconductor Corp. National Semiconductor CMOS Databook. 1981. p. 3-63.
Fig. 4-8: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-240.
Fig. 4-9: Teledyne Semiconductor, Data & Design Manual, 1981, p. 7-39.
Fig. 4-10: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-50.
Fig. 4-11: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-32.
Fig. 4-12: Precision Monolithics Incorporated 1981 Full Line Catalog, p. 8-13.

Chapter 5

Fig. 5-1: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-22.
Fig. 5-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.
Fig. 5-3: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 8-64.
Fig. 5-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 12-39.

Chapter 6

Fig. 6-1: Electronics Today International, 3/82, p. 66.
Fig. 6-2: 101 Electronic Projects, 1977, IC 23.
Fig. 6-3: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-66.
Fig. 6-4: Electronics Today International, 10/79, p. 93.
Fig. 6-5: No reference.
Fig. 6-6: No reference.
Fig. 6-7: Electronics Today International, 3/75, p. 66.
Fig. 6-8: Electronics Today International, 3/78, p. 52.
Fig. 6-9: Electronics Today International, 5/78, p. 85.
Fig. 6-10: Modern Electronics, 7/78, p. 58.

Chapter 7

Fig. 7-1: Courtesy of Fairchild Camera & Instrument Corporation. Fairchild Semiconductor Application Note 300.
Fig. 7-2: Ham Radio, 1/78, p. 78.
Fig. 7-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-23.
Fig. 7-4: 73 Magazine, 12/76, p. 97.
Fig. 7-5: 73 Magazine, 7/77, p. 34.
Fig. 7-6: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN29-9.
Fig. 7-7: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, P. LB16-1.
Fig. 7-8: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-31.
Fig. 7-9: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-25.
Fig. 7-10: How to Design/Build Remote Control Devices TAB Book No. 1277, p. 230.
Fig. 7-11: Radio Electronics, 7/83, p. 7.
Fig. 7-12: Electronics Today International, Summer 1982, p. 45.
Fig. 7-13: 73 Magazine, p. 31.
Fig. 7-14: Reprinted from Electronics, 11/83. Copyright 1983, McGraw Hill Inc. All rights reserved.
Fig. 7-15: Electronics Today International, 7/72, p. 84.
Fig. 7-16: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-42.
Fig. 7-17: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-171.

Chapter 8

Fig. 8-1: Courtesy of Fairchild Camera & Instrument Corporation, Fairchild Progress, 11-12/76, p. 26.
Fig. 8-2: Courtesy of Fairchild Camera & Instrument Corporation. Fairchild Progress, 5-6/77, p. 22.
Fig. 8-3: Reprinted with the permission

of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-44.
Fig. 8-4: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-14.
Fig. 8-5: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-14.
Fig. 8-6: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-23.
Fig. 8-7: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-51. Application Note AN125, p. 7.
Fig. 8-8: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-51. Application Note AN125, p. 6.
Fig. 8-9: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-171.
Fig. 8-10: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-63.
Fig. 8-11: No reference.
Fig. 8-12: Electronics Today International, 3/78, p. 81.
Fig. 8-13: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B, p. 8-21.
Fig. 8-14: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B, p. 8-21.
Fig. 8-15: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B, p. 8-21.
Fig. 8-16: Reprinted with the permission of National Semiconductor Corp. National Semiconductor Application Note AN125, p. 7.
Fig. 8-17: Reprinted with the permission of National Semiconductor Corp. Application Note AN69, p. 4.
Fig. 8-18: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-25.
Fig. 8-19: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-17.
Fig. 8-20: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-170.
Fig. 8-21: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-170.
Fig. 8-22: Reprinted with permission of National Semiconductor, Corp. Application Note AN69, p. 4.
Fig. 8-23: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-89.
Fig. 8-24: Reprinted with permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-203.

Chapter 9

Fig. 9-1: Canadian Projects Number 1, Spring/78, p. 27.
Fig. 9-2: No reference.
Fig. 9-3: Electronics Today International, 4/79, p. 18.
Fig. 9-4: Reprinted with permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-389.
Fig. 9-5: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.
Fig. 9-6: Reprinted with permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-91.
Fig. 9-7: Reprinted with permision of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-45.
Fig. 9-8: Reprinted with permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-43.
Fig. 9-9: Reprinted with permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-28.
Fig. 9-10: Signetics Analog Data Manual, 1982, p. 4-8.
Fig. 9-11: Signetics Analog Data Manual, 1982, p. 15-6.
Fig. 9-12: Signetics Analog Data Manual, 1977, p. 466.
Fig. 9-13: Reprinted with permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-27.
Fig. 9-14: Reprinted with permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-32.
Fig. 9-15: Signetics Analog Data Manual, 1982, p. 15-6.
Fig. 9-16: Signetics Analog Data Manual, 1977, p. 466.
Fig. 9-17: Reprinted with permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-88.
Fig. 9-18: Reprinted with permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-20.
Fig. 9-19: Reprinted with permission of National Semiconductor Corp. Audio/Radio Handbook, 1980. p. 2-21.
Fig. 9-20: Signetics Analog Data Manual, 1977, p. 466.
Fig. 9-21: Signetics Analog Data Manual, 1983, p. 10-92.
Fig. 9-22: Signetics Analog Data Manual, 1982, p. 15-6.

Chapter 10

Fig. 10-1: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN162-10.
Fig. 10-2: Electronics Today International, 6/79, p. 75.
Fig. 10-3: Signetics 555 Timers, 1973, p. 24.
Fig. 10-4: Electronics Today International, 12/75, p. 72.
Fig. 10-5: Electronics Today International, 2/75, p. 51.
Fig. 10-6: Electronics Today International, 7/81, p. 22.
Fig. 10-7: Electronics Today International, 7/77, p. 32.
Fig. 10-8: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. LB33-1.
Fig. 10-9: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-141.
Fig. 10-10: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-138.
Fig. 10-11: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-31.
Fig. 10-12: 73 Magazine, 7/77, p. 34.
Fig. 10-13: Modern Electronics, 2/78, p. 56.
Fig. 10-14: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-140.
Fig. 10-15: The Build-It Book Of Electronic Projects, TAB Book No. 1498, p. 80.
Fig. 10-16: 73 Magazine, 1/82, p. 41.
Fig. 10-17: Electronics Today International, 10/77, p. 47.
Fig. 10-18: Modern Electronics, 9/78, p. 37.
Fig. 10-19: Electronics Today International, 10/77, p. 38.
Fig. 10-20: The Build-It Book Of Electronic Projects, TAB Book No. 1498, p. 111.
Fig. 10-21: Modern Electronics, 5/78, p. 7.
Fig. 10-22: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-143.
Fig. 10-23: Reprinted with the permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 207.

Fig. 10-24: No reference.

Chapter 11

Fig. 11-1: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 7-32.
Fig. 11-2: 101 Electronics Projects, 1977, p. 97.
Fig. 11-3: Courtesy of Motorola Inc. Application Note AN-294, p. 6.
Fig. 11-4: 73 Magazine, 2/79, p. 156.
Fig. 11-5: 73 Magazine, 7/77.
Fig. 11-6: Ham Radio, 12/79, p. 67.
Fig. 11-7: 73 Magazine, 2/83, p. 99.
Fig. 11-8: 44 Electronics Projects For SWLS, CBers & Radio Experimenters, TAB Book No. 1258, p. 153.
Fig. 11-9: Yuasa Battery (America) Inc. Application Manual for NP type battery.
Fig. 11-10: Electronics Today International, 11/80.
Fig. 11-11: 73 Magazine, 7/77.
Fig. 11-12: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 203.
Fig. 11-14: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-31.
Fig. 11-15: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-141.

Chapter 12

Fig. 12-1: NASA Tech Brief, B73-10249.
Fig. 12-2:Electronics Today International, 1/75, p. 66.
Fig. 12-3: Electronics Australia, 2/76, p. 91.
Fig. 12-4: 73 Magazine, 2/79, p. 78.
Fig. 12-5: Electronics Today International, 6/79, p. 103.
Fig. 12-6: Ham Radio, 9/82, p. 78.
Fig. 12-7: Courtesy of Texas Instruments Incorporated. Optoelectronics Databook, 1983-84, p. 15-5.
Fig. 12-8: 73 Magazine, 2/79, p. 78.
Fig. 12-9: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-19.
Fig. 12-10: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-109.
Fig. 12-11: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-109.

Chapter 13

Fig. 13-1: Intersil Data Book, 5/83, p. 5-238.
Fig. 13-2: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-131.
Fig. 13-3: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-160.
Fig. 13-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-17.
Fig. 13-5: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-31.
Fig. 13-6: Precision Monolithics Incorporated, 1981 Full Line Catalog. p. 16-159.
Fig. 13-7: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-324.
Fig. 13-8: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-324.
Fig. 13-9: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-35.
Fig. 13-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-11.

Chapter 14

Fig. 14-1: Radio - Electronics, 1/67.
Fig. 14-2: Modern Electronics, 2/78, p. 17.
Fig. 14-3: Electronics Today International, 5/75, p. 68.
Fig. 14-4: Electronics Today International, 4/78, p. 81.
Fig. 14-5: Modern Electronics, 6/78, p. 14.
Fig. 14-6: Reprinted with permission from General Electric Semiconductor Department. General Electric, 2/68.
Fig. 14-7: Electronics Today International, 6/74, p. 67.
Fig. 14-8: Modern Electronics, 2/78, p. 16.
Fig. 14-9: © Siliconix incorporated. T100/T300 Applications.
Fig. 14-10: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 224.
Fig. 14-11: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-143.
Fig. 14-12: Electronics Today International, 6/82, p. 69.
Fig. 14-13: ©Siliconix incorporated. Siliconix Application Note AN154.
Fig. 14-14: Wireless World, 5/78, p. 69.
Fig. 14-15: Reprinted with permission from General Electric Semiconductor Department. General Electric, 2/68.

Chapter 15

Fig. 15-1. Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN146-1.
Fig. 15-2: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-112.
Fig. 15-3: Supertex Data Book, 1983, p. 5-23.
Fig. 15-4: Supertex Data Book, 1983, p. 5-22.
Fig. 15-5: How To Design/Build Remote Control Devices, TAB Book No. 1277, p. 287.
Fig. 15-6: How To Design/Build Remote Control Devices, TAB Book No. 1277, p. 289.
Fig. 15-7: How To Design/Build Remote Control Devices, TAB Book No. 1277, p. 290.
Fig. 15-8: How To Design/Build Remote Control Devices, TAB Book No. 1277, p. 291.
Fig. 15-9: Signetics Analog Data Manual, 1982, p. 16-28.

Chapter 16

Fig. 16-1: Reprinted from Electronics, 6/78, p. 150. Copyright 1978, McGraw Hill Inc. All rights reserved.
Fig. 16-2: Reprinted from Electronics, 5/73, p. 96. Copyright 1973, McGraw Hill Inc. All rights reserved.
Fig. 16-3: 303 Dynamic Electronic Circuits, TAB Book No. 1060, p. 290.
Fig. 16-4: 73 Magazine, 2/79, p. 79.
Fig. 16-5: Wireless World, 12/74, p. 504.
Fig. 16-6: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-123.
Fig. 16-7: Electronics Today International, 3/78, p. 51.
Fig. 16-8: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-215.
Fig. 16-9: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-17.
Fig. 16-10: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, 1979, p. 7-8.
Fig. 16-11: Courtesy of Motorola Inc. Linear Interface Circuits, 1979, p. 7-8.
Fig. 16-12: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-123.

Fig. 16-13: Siliconix Application Note AN73-6, p. 5.
Fig. 16-14: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-31.
Fig. 16-15: Precision Monolithics Incorporated 1981 Fall Line Catalog, p. 8-31.
Fig. 16-16: Teledyne Semiconductor, Databook, p. 9.
Fig. 16-17: ©Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-4.
Fig. 16-18: Signetics Analog Data Manual, 1982, p. 8-14.
Fig. 16-19: Precision Monolithics Incorporated 1981 Full Line Catalog, p. 8-12.
Fig. 16-20: Signetics Analog Data Manual, 1982, p. 3-38.
Fig. 16-21: Harris Semiconductor, Linear & Data Acquisition Products, p. 2-46.
Fig. 16-22: Harris Semiconductor Application Note 509.

Chapter 17

Fig. 17-1: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN240-5.
Fig. 17-2: Electronics Today International, 10/77, p. 45.
Fig. 17-3: ©Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 7-29.
Fig. 17-4: Reprinted with the permission of National Semiconductor Corp. National Semiconductor CMOS Databook, 1981, p. 3-61.
Fig. 17-5: Precision Monolithics Incorporated 1981 Full Line Catalog, p. 16-142.
Fig. 17-6: ™Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 7-29.
Fig. 17-7: Electronics Today International, 10/77, p. 39
Fig. 17-8: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-28.
Fig. 17-9: ©Siliconix Incorporated. T100/T300 Applications.
Fig. 17-10: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN240-2.
Fig. 17-11: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 7-30.
Fig. 17-12: Signetics Analog Data Manual, 1982, p. 3-71.
Fig. 17-13: Signetics Analog Data Manual, 1982, p. 6-20.
Fig. 17-14: Signetics Analog Data Manual, 1983, p. 10-99.
Fig. 17-15: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 6-27.
Fig. 17-16: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 8-258.
Fig. 17-17: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-50.
Fig. 17-18: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 8-258.
Fig. 17-19: ©Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 7-31.
Fig. 17-20: Signetics Analog Data Manual, 1982, p. 3-15.
Fig. 17-21: RCA Corporation, Solid State Division, Digital Integrated Circuits Application Note ICAN-6346, p. 4.
Fig. 17-22: ©Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 6-42.
Fig. 17-23: Signetics Analog Data Manual, 1982, p. 8-14.
Fig. 17-24: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-56.

Chapter 18

Fig. 18-1: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-4.
Fig. 18-2: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-5.
Fig. 18-3: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-4.

Chapter 19

Fig. 19-1: Courtesy of Motorola Inc. Application Note AN-417B, p. 5.
Fig. 19-2: Courtesy of Motorola Inc. Application Note AN417B, p. 3.
Fig. 19-3: The Complete Handbook of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 326.
Fig. 19-4: Electronics Today International, 1/76, p. 46.
Fig. 19-5: Ham Radio, 2/79, p. 40.
Fig. 19-6. Electronics Today International, 8/83, p. 57.
Fig. 19-7: Electronics Today International, 11/76, p. 44.
Fig. 19-8: Ham Radio, 2/79, p. 40.
Fig. 19-9: Ham Radio, 2/79, p. 42.
Fig. 19-10: Ham Radio, 2/79, p. 41.
Fig. 19-11: Ham Radio, 2/79, p. 43.
Fig. 19-12: Ham Radio, 2/79, p. 43.
Fig. 19-13: Ham Radio, 2/79, p. 43.
Fig. 19-14: Ham Radio, 2/79, p. 43.
Fig. 19-15: Ham Radio, 2/79, p. 38.
Fig. 19-16: Ham Radio, 2/79, p. 39.
Fig. 19-17: Ham Radio, 3/82, p. 66.
Fig. 19-18: Electronics Today International, 8/73, p. 82.
Fig. 19-19: The Complete Handbook of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 322.
Fig. 19-20: Ham Radio, 4/78, p. 51.
Fig. 19-21: Modern Electronics, 6/78, p. 57.
Fig. 19-22: The Complete Handbook of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 336.
Fig. 19-23: 73 Magazine, 8/78, p. 80.
Fig. 19-24: Third Book Of Electronic Projects, TAB Book No. 1446, p. 22.
Fig. 19-25: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 69.
Fig. 19-26: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 64.
Fig. 19-27: Ham Radio, 4/78, p. 50.
Fig. 19-28: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of
Fig. 19-29: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 63.
Fig. 19-30: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 63.
Fig. 19-31: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 63.
Fig. 19-32: CHRYSTAL OSCIL-

LATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 63.

Fig. 19-33: Third Book Of Electronic Projects, TAB Book No. 1446, p. 21.

Fig. 19-34: Intersil.

Fig. 19-35: The Complete Handbook Of Amplifiers, Oscillators & Multivibrators, Tab Book No. 1230, p. 324.

Fig. 19-36: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 64.

Fig. 19-37: The Complete Handbook Of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 325.

Fig. 19-38: Ham Radio, 2/79, p. 41.

Fig. 19-40. The Complete Handbook Of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 330.

Fig. 19-41: The Complete Handbook Of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 331.

Fig. 19-42: Ham Radio, 4/78, p. 50.

Fig. 19-43: Ham Radio, 2/79, p. 40.

Fig. 19-44: 73 Magazine.

Fig. 19-45: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-241.

Fig. 19-46: Teledyne Semiconductor Databook, p. 9.

Fig. 19-47: Reprinted with the permission of National Semiconductor Corp. Application Note 32, p. 8.

Fig. 19-48: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-26.

Fig. 19-49: Ham Radio, 2/79, p. 40.

Fig. 19-50: CHRYSTAL OSCILLATOR CIRCUITS, Robert J. Matthys, Copyright © 1983, John Wiley & Sons, Inc. Reprinted by permission of John Wiley & Sons, Inc. r.f. Design, 5-6/83, p. 66.

Chapter 20

Fig. 20-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-123.

Fig. 20-2: Intersil Data Book, 5/83, p. 5-289.

Fig. 20-3: Reprinted with the permission of National Semiconductor Corp. Application Note AN-71, p. 5.

Fig. 20-4: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 305.

Fig. 20-5: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-35.

Chapter 21

Fig. 21-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-123.

Fig. 21-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-30.

Fig. 21-3: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-112.

Fig. 21-4: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-30.

Chapter 22

Fig. 22-1: Electronics Today International, 9/75, p. 65.

Fig. 22-2: Signetics Analog Data Manual, 1982, p. 6-13.

Fig. 22-3: Electronic Today International, 8/79, p. 99.

Fig. 22-4: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-15.

Fig. 22-5: © Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 6-41.

Fig. 22-6: Signetics Analog Data Manual, 1982, p. 6-21.

Fig. 22-7: Signetics Analog Data Manual, 1982, p. 6-21.

Chapter 23

Fig. 23-1: Ham Radio 11/78, p. 64.

Fig. 23-2: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 2-5.

Fig. 23-3: Signetics Analog Data Manual, 1983, p. 11-15.

Fig. 23-4: Signetics Analog Data Manual, 1983, p. 11-10.

Fig. 23-5: Signetics Analog Data Manual, 1982, p. 16-28.

Fig. 23-6: Signetics Analog Manual, 1982, p. 16-28.

Chapter 24

Fig. 24-1: Signetics 555 Timers, 1973, p. 19.

Fig. 24-2: Courtesy of Motorola Inc: Linear Interface Integrated Circuits, 1979, p. 7-30.

Fig. 24-3: Electronics Today International, 1/76, p. 45.

Fig. 24-4: Precision Monolithics Incorporated 1981 Full Line Catalog, p., 8-33.

Fig. 24-5: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 219.

Fig. 24-6: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 218.

Fig. 24-7: Courtesy of Motorola Inc. Application Note AN294.

Fig. 24-8: Signetics 555 Timers, 1973, p. 20.

Chapter 25

Fig. 25-1: Radio-Electronics, 2/83, p. 76.

Fig. 25-2: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-98.

Fig. 25-3: Radio-Electronics, 12/78, p. 77.

Fig. 25-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 14-17.

Fig. 25-5: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 14-17.

Fig. 25-6: Electronics Today International, 3/78, p. 50.

Fig. 25-7: RCA Corp., Solid State Division, Digital Integrated Circuits Application Note ICAN-6346, p. 5.

Fig. 25-8: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-97.

Fig. 25-9: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-25.

Fig. 25-10: Reprinted with the permission of National Semiconductor Corp. National Semiconductor, Application Note LB-25.

Fig. 25-11: Electronics Today International, 9/72, p. 86.

Fig. 25-12: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 56.

Fig. 25-13: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-180.

Fig. 25-14: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-9.

Fig. 25-15: Signetics Analog Data Manual, 1983, p. 10-100.

Fig. 25-16: © Siliconix incorporated. Siliconix Application Note AN73-6, p. 4.

Fig. 25-17: Signetics Analog Data Manual, 1983, p. 13-6.
Fig. 25-18: Signetics 555 Timers, 1973, p. 17.
Fig. 25-19: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-123.
Fig. 25-20: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 205.
Fig. 25-21: Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-14.
Fig. 25-22: Signetics Analog Data Manual, 1983, p. 11-9.
Fig. 25-23: Signetics Analog Data Manual, 1983, p. 11-9.
Fig. 25-24:Signetics Analog Data Manual, 1983, p. 10-100.
Fig. 25-25: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-38.
Fig. 25-26: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-12.
Fig. 25-27: Signetics Analog Data Manual, 1977, p. 264.
Fig. 25-28: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-31.
Fig. 25-29: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-38.

Chapter 26

Fig. 26-1: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 8-5.
Fig, 26-2: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 8-4.
Fig. 26-3: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-10.
Fig. 26-4. Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 11-55.
Fig. 26-5: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-10.
Fig. 26-6: Ferranti, Technical Handbook Vol. 10, Data Converters, 1983, p. 1-25.
Fig. 26-7: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-50.
Fig. 26-8: ©Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 8-5.
Fig. 26-9. Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-7.
Fig. 26-10. Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 11-55.
Fig. 26-11: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 8-20.
Fig. 26-12:Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-17.
Fig. 26-13: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-50.
Fig. 26-14: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 11-54.
Fig. 26-15: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-159.

Chapter 27

Fig. 27-1: Ham Radio, 8/81, p. 27.
Fig. 27-2: Ham Radio, 8/81, p. 28.
Fig. 27-3: Ham Radio, 8/81, p. 27.
Fig. 27-4: Ham Radio, 8/81, p. 26.
Fig. 27-5: Ham Radio, 8/81, p. 26.
Fig. 27-6: Ham Radio, 6/77, p. 42.
Fig. 27-7: Ham Radio, 8/81, p. 27.

Chapter 28

Fig. 28-1: Reprinted from Electronics, 12/74. p. 105. Copyright 1974, Mc-Graw Hill Inc. All rights reserved.
Fig. 28-2: Electronics Today International, 10/82, p. 80.
Fig. 28-3: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-188.
Fig. 28-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-172.
Fig. 28-5: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, 1979, p. 5-102.
Fig. 28-6: Intersil Data Book, 5/83, p. 6-52.
Fig. 28-7: Reprinted with the permission of National Semiconductor Corp., Linear Databook, 1982, p. 9-171.
Fig. 28-8: Electronics Today International, 3/78, p. 50.
Fig. 28-9: Intersil Data Book, 5/83, p. 6-34.

Chapter 29

Fig. 29-1: Ham Radio, 1/78, p. 94
Fig. 29-2: Reprinted with permission from General Electric Semiconductor Department GE Semiconductor Data Handbook, Third Edition, p. 577.
Fig. 29-3: Reprinted with permission from General Electric Semiconductor Department GE Semiconductor Data Handbook, Third Edition, p. 577.
Fig. 29-4: Reprinted with permission from General Electric Semiconductor Department GE Semiconductor Data Handbook, Third Edition, p. 573.
Fig. 29-5: Reprinted with permission from General Electric Semiconductor Department GE Semiconductor Data Handbook, Third Edition, p. 183.

Chapter 30

Fig. 30-1: Reprinted with the permission of National Semiconductor Corp. National Semiconductor CMOS Databook, 1981, p. 8-44.
Fig. 30-2: Electronics Today International, 4/79, p. 22.
Fig. 30-3: SGS-ATES Databook COS/MOS B-Series, 2/82, p. 548.
Fig. 30-4: ©Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 6-60.
Fig. 30-5: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-81.
Fig. 30-6: Signetics Analog Data Manual, 1982, p. 8-10.
Fig. 30-7. Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-19.
Fig. 30-8.Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-159.
Fig. 30-9. Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-159.
Fig. 30-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-11.
Fig. 30-11: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-21.
Fig. 30-12: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-167.
Fig. 30-13: SGS-ATES Databook COS/MOS B-Series, 2/82, p. 548.
Fig. 30-14: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-123.
Fig. 30-15: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-123.
Fig. 30-16: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-39.
Fig. 30-17: SGS-ATES Databook COS/MOS B-Series, 2/82, p. 548.

Chapter 31

Fig. 31-1: Reprinted with permission from General Electric Semiconductor Department. Optoelectronics, Second Edition, p. 113.
Fig. 31-2: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 13-11.
Fig. 31-3: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-153.
Fig. 31-4: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 13-14.
Fig. 31-5: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 13-20.
Fig. 31-6: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 13-20.

Chapter 32

Fig. 32-1: No reference.
Fig. 32-2: No reference.
Fig. 32-3: Modern Electronics, 2/78, p. 47.
Fig. 32-4: No reference.
Fig. 32-5: The Giant Book Of Electronics Projects, TAB Book No. 1367, p. 480.
Fig. 32-6: The Giant Book Of Electronics Projects, TAB Book No. 1367, p. 114.
Fig. 32-7: The Giant Book Of Electronics Projects, TAB Book No. 1367, p. 114
Fig. 32-8: 73 Magazine.

Chapter 33

Fig. 33-1: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-58.
Fig. 33-2: Intersil Data Book, 5/83, p. 3-135.
Fig. 33-3: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-114.
Fig. 33-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-50.
Fig. 33-5: Electronics, 9/76, p. 100.
Fig. 33-6: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquision Databook, 1980, p. 3-117.
Fig. 33-7: Reprinted from Electronics, 12/78, p. 124. Copyright 1978, Mc-Graw Hill Inc. All rights reserved.
Fig. 33-8: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, 17-132.
Fig. 33-9: Reprinted with the permission of National Semiconductor Corp. Application Note LB-5, p. 1.
Fig. 33-10. Electronics Today International, 11/74, p. 67.
Fig. 33-11. Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-180.
Fig. 33-12. Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-179.
Fig. 33-13. Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-41.
Fig. 33-14. Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-119.
Fig. 33-15: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-177.
Fig. 33-16: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-178.
Fig. 33-17: 73 Magazine, 4/79, p. 42.
Fig. 33-18: 303 Dynamic Electronic Circuits, TAB Book No. 1060, p. 289.
Fig. 33-19: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-15.
Fig. 33-20:Signetics Analog Data Manual, 1982, p. 3-77.
Fig. 33-21: Harris Semiconductor, Linear & Data Acquisition Products, p. 2-85.
Fig. 33-22: © Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-9.
Fig. 33-23: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-104.
Fig. 33-24: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1o80, p. 3-23.
Fig. 33-25: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-116.
Fig. 33-26; Signetics Analog Data Manual, 1982, p. 4-8.
Fig. 33-27: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-115.
Fig. 33-28: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-116.
Fig. 33-29: Harris Semiconductor, Linear & Data Acquisition Products, p. 2-84.
Fig. 33-30: Courtesy of Motorola Inc. Motorola Semiconductor Library Vol. 6, Series B, p. 3-126.
Fig. 33-31: Ham Radio, 2/78, p. 72.
Fig. 33-32: Signetics Analog Data Manual, p. 401.
Fig. 33-33: Signetics Analog Data Manual, p. 75.
Fig. 33-34: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-58.
Fig. 33-35: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-97.
Fig. 33-36: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-157.
Fig. 33-37: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-11.
Fig. 33-38: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-158.
Fig. 33-39: 73 Magazine, 1/79, p. 127.
Fig. 33-40: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-131.
Fig. 33-41: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-59.
Fig. 33-42: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-56.
Fig. 33-43:Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-58.

Chapter 34

Fig. 34-1: Reprinted with permission from General Electric Semiconductor Department. GE Application Note 201.10.
Fig. 34-2: Electronics Today International, 4/75, p. 42.
Fig. 34-3: © Siliconix incorporated, Application Note AN154.
Fig. 34-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-289.
Fig. 34-5: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Second Edition, p. 905.
Fig. 34-6: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 573.
Fig. 34-7: Radio-Electronics, 5/79, p. 84.

Fig. 34-8: 49 Easy To Build Electronic Projects, TAB Book No. 1337, p. 22.
Fig. 34-9: 49 Easy To Build Electronic Projects, TAB Book No. 1337, p. 98.
Fig. 34-10: Electronics Today International, 12/74, p. 66.
Fig. 34-11: No reference.
Fig. 34-12: Electronics Today International, 5-75, p. 67.
Fig. 34-13: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 205.
Fig. 34-14: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 207.
Fig. 34-15: Reprinted vith the permission of National Semiconductor Corp. Linear Databook, 1982, p. 12-14.
Fig. 34-16: © Siliconix incorporated, Application Note AN154.
Fig. 34-17: © Siliconix incorporated, Application Note AN154.
Fig. 34-18: © Siliconix incorporated, Application Note AN154.
Fig. 34-19: 0Siliconix incorporated, Application Note AN154.
Fig. 34-20: © Siliconix incorporated, Application Note AN154.
Fig. 34-21: © Siliconix incorporated, Application Note AN154.
Fig. 34-22: © Siliconix incorporated, Application Note AN154.
Fig. 34-23: © Siliconix incorporated, Application Note AN154.
Fig. 34-24: © Siliconix incorporated, Application Note AN154.
Fig. 34-25: © Siliconix incorporated, Application Note AN154.
Fig. 34-26: © Siliconix incorporated, Application Note AN154.

Chapter 35

Fig. 35-1: Intersil Data Book, 5/83, p. 6-49.
Fig. 35-2: The Giant Book Of Electronic Projects, TAB Book No. 1367, p. 109.
Fig. 35-3: 73 Magazine, 6/83, p. 106.
Fig. 35-4: 104 Weekend Electronic Projects, TAB Book No. 1436, p. 166.

Chapter 36

Fig. 36-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-110.
Fig. 36-2: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 5-9.
Fig. 36-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-99.
Fig. 36-4: Courtesy of Motorola Inc. Linear Integrated Circuits, p. 6-99.
Fig. 36-5: Signetics Analog Data Manual, 1982, p. 16-29.

Chapter 37

Fig. 37-1: Teledyne Semiconductor Publication DG-114-87, p. 7.
Fig. 37-2: ©Siliconix incorporated, Analog Switch & IC Product Data Book, 1/82, p. 7-30.
Fig. 37-3: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-140.
Fig. 37-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 8-257.
Fig. 37-5: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 12-20.
Fig. 37-6: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-143.
Fig. 37-7: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 8-257.

Chapter 38

Fig. 38-1: Electronics Today International, 1/77, p. 83.
Fig. 38-2: 101 Electronic Projects, 1975, #32.
Fig. 38-3: Electronics Today International, 10/76, p. 66.
Fig. 38-4: Electronics Today International, 4/75, p. 67.
Fig. 38-5: Canadian Project Number 1, Spring 78, p. 55.
Fig. 38-6: Electronics Today International, 11/76, p. 44.

Chapter 39

Fig. 39-1: Modern Electronics, 2/78, p. 49.
Fig. 39-2: Electronics Today International, 10/78, p. 103.
Fig. 39-3: Radio-Electronics, 3/78, p. 76.
Fig. 39-4: Popular Mechanics, 5/78, p. 45.
Fig. 39-5: 303 Dynamic Electronic Circuits, TAB Book No. 1060, p. 36.
Fig. 39-6: Electronics Today International, 9/82, p. 70.
Fig. 39-7: Electronics Today International, 4/78, p. 77.
Fig. 39-8: 73 Magazine.
Fig. 39-9: No reference
Fig. 39-10: Electronics Today International, 2/77, p. 73.

Chapter 40

Fig. 40-1: Reprinted with permission of Control Engineering, 1301 S. Grove Ave. Barrington, Illinois 12/73, p. 43.
Fig. 40-2: Courtesy of Motorola Inc. Communications Engineering Bulletin EB-33.
Fig. 40-3: Courtesy of Motorola Inc. Communications Engineering Bulletin EB-33.

Chapter 41

Fig. 41-1: Courtesy of Texas Instruments Incorporated. Optoelectronics Databook, 1983-84, p. 15-12.
Fig. 41-2: 73 Magazine, 7/77, p. 35.
Fig. 41-3: Electronics Today International, 6/76, p. 40.
Fig. 41-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-172.
Fig. 41-5: Courtesy of Texas Instruments Incorporated. Optoelectronics Databook, 1983-84, p. 15-11.
Fig. 41-6: Signetics Analog Data Manual, 1982, p. 8-14.
Fig. 41-7: ©Siliconix incorporated, Analog Switch & IC Product Data Book, 1/82, p. 6-14.
Fig. 41-8: 73 Magazine.
Fig. 41-9: Reprinted from Electronics, 3/73, p. 119. Copyright 1973, McGraw Hill Inc. All rights reserved.

Chapter 42

Fig. 42-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-127.
Fig. 42-2: Supertex Data Book, 1983, p. 5-20.
Fig. 42-3: Plessey Semiconductors, Linear IC Handbook, 5/82, p. 86.
Fig. 42-4: Plessey Semiconductors, Linear IC Handbook, 5/82, p. 91.
Fig. 42-5: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-74.
Fig. 42-6: Electronics Today International, 6/82, p. 70.

Chapter 43

Fig. 43-1: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 2-85.
Fig. 43-2: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-77.
Fig. 43-3: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-178.
Fig. 43-4: Courtesy of Fairchild Cam-

era & Instrument Corporation. Linear Databook, 1982, p. 4-43.
Fig. 43-5: Reprinted with the permission of National Semiconductor Corp. Application Note 32, p. 5.
Fig. 43-6: Reprinted with the permission of National Semiconductor Corp. Application Note LB1, p. 2.
Fig. 43-7: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 120.
Fig. 43-8: ©Siliconix incorporated. T100/T300 Applications.
Fig. 43-9: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 4-27.
Fig. 43-10: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN242-15.
Fig. 43-11: Signetics Analog Data Manual, 1982, p. 3-71.
Fig. 43-12: ©Siliconix incorporated. Application Note, AN73-6, p. 3.
Fig. 43-13: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 3-7.
Fig. 43-14: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-102.
Fig. 43-15: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-82.
Fig. 43-16: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-171.
Fig. 43-17: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 122.
Fig. 43-18: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-11.
Fig. 43-19: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-6.
Fig. 43-20: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-159.
Fig. 43-21: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-56.
Fig. 43-22: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-92.
Fig. 43-23: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-50.
Fig. 43-24: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-37.
Fig. 43-25: Signetics Analog Data Manual, 1982, p. 3-15.

Chapter 44

Fig. 44-1: Courtesy of Texas Instruments Incorporated. Optoelectronics Databook, 1983, p. 15-13.
Fig. 44-2: CQ, 3/78, p. 72.
Fig. 44-3: Signetics Analog Data Manual, 1982, p. 3-76.
Fig. 44-4: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 207.
Fig. 44-5: Reprinted with permission from General Electric Semiconductor Department. General Electric Newsletter, Vol. 11. No. 1, p. 5.
Fig. 44-6: Reprinted with permission from General Electric Semiconductor Department. Optoelectronics, Second Edition, p. 112.
Fig. 44-7: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-42.
Fig. 44-8: Electronics Today International, 5/77, p. 77.
Fig. 44-9: Reprinted from Computers & Electronics. Copyright Ziff-Davis Publishing Company. 4/83, p. 109.
Fig. 44-10: The Build-It Book Of Electronic Projects, TAB Book No. 1498, p. 42.
Fig. 44-11: Copyright by Computer Design. All rights reserved. Reprinted by permission. 1/83, p. 77.
Fig. 44-12: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 440.
Fig. 44-13: Copyright by Computer Design. All rights reserved. Reprinted by permission. 1/83, p. 77.
Fig. 44-14: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 1371-4.
Fig. 44-15: Precision Monolithics Incorporated, Linear & Conversion IC Products, 7/78, p. 7-12.
Fig. 44-16: Electronic Projects, 1977, p. 82.
Fig. 44-17: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-109.
Fig. 44-18: Reprinted with permission from General Electric Semiconductor Department. Optoelectronics, Second Edition, p. 111
Fig. 44-19: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-88.

Chapter 45

Fig. 45-1: RCA Corporation, RCA Solid-State Devices Manual, 1975, p. 734.
Fig. 45-2: Reprinted with permission from General Electric Semiconductor Department. GE Project H5, p. 157.
Fig. 45-3: Solid State Products, New Design Idea, No. 5.
Fig. 45-4: Reprinted from Electronics, 12/74, p. 111. Copyright 1974, McGraw Hill Inc. All rights reserved.
Fig. 45-5: Electronics Today International, 12/72, p. 86.
Fig. 45-6: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook. Second Edition, p. 585.
Fig. 45-7: 101 Electronic Projects, 1975.
Fig. 45-8: Courtesy of Motorola Inc. Motorola Semiconductor Products. Circuit Applications for the Triac (AN-466), p. 12.
Fig. 45-9: Courtesy of Motorola Inc. Motorla Semiconductor Products Circuit Applications for the Triac (AN-466), p. 5.
Fig. 45-10: Electronics Today International, 7/75, p. 41.
Fig. 45-11: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual Sixth Edition, 1979, p. 264.
Fig. 45-12: Courtesy of Motorola Inc. Motorola Semiconductor Products Circuit Applications for the Triac (AN-466), p. 6.
Fig. 45-13: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 443.
Fig. 45-14: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual Sixth Edition, 1979, p. 114.
Fig. 45-15: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition. p. 64.
Fig. 45-16: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Second Edition, p. 727.
Fig. 45-17: Solid State Products, New Design Idea, No. 9.
Fig. 45-18: Reprinted with the permis-

sion of National Semiconductor Corp. Transistor Databook, 1982, p. 7-35.
Fig. 45-19: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Second Edition, p. 727.
Fig. 45-20: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-111.
Fig. 45-21: SGS-ATES Databook COS/MOS B-Series, 2/82, p. 548.

Chapter 46

Fig. 46-1: Machine Design, 9/80, p. 126.
Fig. 46-2: Machine Design, 9/80, p. 127.
Fig. 46-3: Reprinted with the permission of National Semiconductor Corp. linear Databook, 1982, p. 9-191.
Fig. 46-4: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-91.
Fig. 46-5: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-89.
Fig. 46-6: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 13-50.

Chapter 47

Fig. 47-1: NASA Tech Briefs, Spring 1983, p. 249.
Fig. 47-2: Courtesy of Texas Instruments Incorporated. Optoelectronics Databook, 1983-84, p. 15-9.
Fig. 47-3: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-93.
Fig. 47-4: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 226.
Fig. 47-5: Modern Electronics, 7/78, p. 55.
Fig. 47-6: Electronics Today International, 8/74, p. 66.
Fig. 47-7: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Application Note, 200.35, p. 14.
Fig. 47-8: Modern Electronics, 3/78, p. 68.
Fig. 47-9: Modern Electronics, 7/78, p. 55.
Fig. 47-10: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-93.

Chapter 48

Fig. 48-1: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 438.
Fig. 48-2: Electronics Today International, 1/78, p. 83.
Fig. 48-3: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.
Fig. 48-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-138.
Fig. 48-5: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-46.
Fig. 48-6: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-48.
Fig. 48-7: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-32.
Fig. 48-8: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-139.
Fig. 48-9: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-46.

Chapter 49

Fig. 49-1: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-49.
Fig. 49-2: Reprinted with the permission of National Semiconductor Corp. National Semiconductor CMOS Databook, 1981, p. 8-124.
Fig. 49-3: Intersil Data Book, 1978.
Fig. 49-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-86.
Fig. 49-5: Radio-Electronics, 10/77, p. 72.
Fig. 49-6: Electronics Today International, 8/78, p. 91.
Fig. 49-7: Third Book Of Electronic Projects, TAB Book No. 1446, p. 40.
Fig. 49-8: Electronics Today International, 8/73, p. 82.
Fig. 49-9: 303 Dynamic Electronic Circuits, TAB Book No. 1060, p. 153.
Fig. 49-10: Electronics Today International, 10/78, p. 97.
Fig. 49-11: Radio-Electronics, 1/80, p. 68.
Fig. 49-12: Signetics Analog Data Manual, 1983, p. 9-40.
Fig. 49-13: Signetics Analog Data Manual, 1983, p. 9-38.
Fig. 49-14: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-187.
Fig. 49-15: Electronics Today International, 1/76, p. 47.
Fig. 49-16: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-140.
Fig. 49-17: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-25.
Fig. 49-18: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 10-8.
Fig. 49-19: Electronics Today International, 7/75, p. 40.

Chapter 50

Fig. 50-1: Reprinted from Electronics, 12/77, p. 78. Copyright 1978, McGraw Hill Inc. All rights reserved.
Fig. 50-2: 101 Electronic Projects, 1977, p. 48.

Chapter 51

Fig. 51-1: ETI Canada, 7/78, p. 46.
Fig. 51-2: The Build-It Book Of Electronic Projects, TAB Book No. 1498, p. 131.
Fig. 51-3: Modern Electronics, 3/78, p. 7.

Chapter 52

Fig. 52-1: Reprinted with the permission of National Semiconductor Corp. Application Note AN69, p. 6.
Fig. 52-2: Courtesy of Texas Instruments Incorporated. Complex Sound Generator, Bulletin No. DL-S 12612, p. 13.
Fig. 52-3: ©Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 6-60.
Fig. 52-4: Signetics Analog Data Manual, 1983, p. 10-99.
Fig. 52-5: Signetics Analog Data Manual, 1983, p. 10-99.
Fig. 52-6: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-157.
Fig. 52-7: ©Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 6-42.
Fig. 52-8: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook. Second Edition, p. 727.
Fig. 52-9: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-37.
Fig. 52-10: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-139.
Fig. 52-11: Electronics Today International, 6/82, p. 64.

Fig. 52-12: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-139.
Fig. 52-13: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-163.
Fig. 52-14: ©Siliconix incorporated. Application Note AN154.
Fig. 52-15: Signetics Analog Data Manual, 1982, p. 3-50.
Fig. 52-16: Signetics Analog Data Manual, 1983, p. 10-20.
Fig. 52-17: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-10.
Fig. 52-18: FERRANTI, Technical Handbook, Vol. 10, Data Converters, 1983, p. 7-26.
Fig. 52-19: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-60.
Fig. 52-20: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-56.
Fig. 52-21: Signetics Analog Data Manual, 1982, p. 4-8.
Fig. 52-22: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-38.

Chapter 53

Fig. 53-1: ©Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 4-24.
Fig. 53-2: ©Siliconix incorporated. Siliconix Analog Switch & IC Product Data Book, 1/82, p. 4-23.
Fig. 53-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-99.
Fig. 53-4: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-178.
Fig. 53-5: Courtesy of Motorola Inc. Motorola Semiconductor Library, Vol. 6, Series B, p. 8-58.
Fig. 53-6: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 4-26.
Fig. 53-7: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-34.

Chapter 54

Fig. 54-1: Modern Electronics, 3/78, p. 6.
Fig. 54-2: 101 Electronic Projects, 1977, p. 25.
Fig. 54-3: 101 Electronic Projects, 1975, p. 53.

Chapter 55

Fig. 55-1: Courtesy of Motorola Inc. Application Note AN-829.
Fig. 55-2: Radio-Electronics, 8/78, p. 41.
Fig. 55-3: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 288.
Fig. 55-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-137.
Fig. 55-5: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-122.
Fig. 55-6: 44 Electronics Projects for Hams, SWLs, CBers, & Radio Experimenters, TAB Book No. 1258, p. 133.
Fig. 55-7: Signetics 555 Timers, 1973, p. 23.
Fig. 55-8: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-17.
Fig. 55-9: Electronics Australia, 4/78, p. 51.
Fig. 55-10: Signetics Analog Data Manual, 1983, p. 11-9.
Fig. 55-11: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 288.
Fig. 55-12: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-98.
Fig. 55-13: Electronics Today International, 8/83, p. 57.
Fig. 55-14: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-81.
Fig. 55-15: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-16.
Fig. 55-16: The Giant Book Of Electronics Projects, TAB Book No. 1367.

Chapter 56

Fig. 56-1: Electronics Today International, 4/78, p. 63.
Fig. 56-2: Modern Electronics, 5/78, p. 6.
Fig. 56-3: Electronics Today International, 8/78, p. 61.
Fig. 56-4: Electronics Today International, 12/78, p. 93.

Chapter 57

Fig. 57-1: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-201.
Fig. 57-2: Reprinted with permission from General Electric Semiconductor Department. Project H13, p. 191.
Fig. 57-3. Courtesy of Motorola Inc. Circuit Applications for the Triac, AN-466, p. 7.
Fig. 57-4: Courtesy of Motorola Inc. AN-443.
Fig. 57-5: Courtesy of Motorola Inc. AN-198.
Fig. 57-6: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 573.
Fig. 57-7: Intersil Data Book, 5/83, p. 5-261.
Fig. 57-8: 101 Electronic Projects, 1977,p.98.
Fig. 57-9: Reprinted with permission from General Electric Semiconductor Department, GE Application Note 201.7.
Fig. 57-10: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, p. 5-145.
Fig. 57-11: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-167.
Fig. 57-12: 101 Electronic Projects, 1975, p. 55.
Fig. 57-13: Electronics Today International. 6/75.
Fig. 57-14: RCA Solid State Devices Manual, 1975, p. 501.
Fig. 57-15: Modern Electronics, 6/78, p. 56.
Fig. 57-16: Reprinted with permission from General Electric Semiconductor Department. GE Project H16, p. 203.
Fig. 57-17: Electronics Today International, 4/75, p. 65.
Fig. 57-18: Courtesy of Motorola Inc. AN-443.
Fig. 57-19: Reprinted with the permission of National Semiconductor Corp. Application Note AN125, p. 9.
Fig. 57-20: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-114.
Fig. 57-21: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 964.
Fig. 57-22: 101 Electronic Projects, 1977, p. 93.
Fig. 57-23: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-114.

Chapter 58

Fig. 58-1: Courtesy of Texas Instru-

ments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 285.
Fig. 58-2: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 286.
Fig. 58-3: RCA Corporation, Solid State Division, Digital Integrated Circuits Application Note, ICAN-6346, p. 5.
Fig. 58-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-154.
Fig. 58-5: Courtesy of Motorola Inc. Linear Integrated Circuits, p. 6-136.
Fig. 58-6: Courtesy of Motorola Inc. Application Note, AN294.
Fig. 58-7: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-47.
Fig. 58-8: Signetics 555 Timers, 1973, p. 22.
Fig. 58-9: Signetics Analog Data Manual, 1983, p. 15-6.
Fig. 58-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-32.
Fig. 58-11: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-46.
Fig. 58-12: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-46.
Fig. 58-13: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 5-7.

Chapter 59

Fig. 59-1: Electronics Today International, 4/76, p. 23.
Fig. 59-2: Popular Electronics, 4/75, p. 87.
Fig. 59-3: Electronics Today International, 4/78, p. 30.
Fig. 59-4: Popular Electronics, 12/76, p. 28.
Fig. 59-5: The Radio Hobbyist's Handbook, TAB Book No. 1346, p. 256.

Chapter 60

Fig. 60-1: Reprinted from Electronics, 7/72, p. 77. Copyright 1972, McGraw Hill Inc. All rights reserved.
Fig. 60-2: Reprinted from Electronics, 10/73, p. 125. Copyright 1973, McGraw Hill Inc. All rights reserved.
Fig. 60-3: 73 Magazine, 12/76, p. 170.
Fig. 60-4: Electronics Today International, 1978.
Fig. 60-6: CQ, 11/83, p. 72.
Fig. 60-7: Electronics Today International, 7/77, p. 77.

Chapter 61

Fig. 61-1: Machine Design, 7/75, p. 39.
Fig. 61-2: Electronics Today International, 4/73, p. 89.
Fig. 61-3: Signetics Analog Data Manual, 1982, p. 16-28.
Fig. 61-4: Teledyne Semiconductor Data & Design Manual, 1981, p. 11-207.
Fig. 61-5: ©Siliconix incorporated, Analog Switch & IC Product Data Book, 1/82, p. 6-4.
Fig. 61-6: Reprinted with the permission of National Semiconductor Corp. Application Note 32, p. 8.

Chapter 62

Fig. 62-1: Electronics Today International, 4/82, p. 39.
Fig. 62-2: Western Digital, Components Handbook, 1983, p. 577.
Fig. 62-3: Modern Electronics, 2/78, p. 72.
Fig. 62-4: Canadian Projects Number 1, Spring 1978, p. 78.
Fig. 62-5: 101 Electronic Projects, 1977, p. 49.
Fig. 62-6: Electronics Today International, 10/74, p. 67.
Fig. 62-8: 44 Electronics Projects For The Darkroom, TAB Book No. 1248, p. 282.
Fig. 62-9: 44 Electronics Projects For The Darkroom, TAB Book No. 1248, p. 284.
Fig. 62-10: Signetics 555 Timers, 1973, p. 23.

Chapter 63

Fig. 63-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-205.
Fig. 63-2: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-191.
Fig. 63-3: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 374.
Fig. 63-4: Reprinted with the permission of National Semiconductor Corp. Application Note 222.
Fig. 63-5: Courtesy of Motorola Inc. Motorola Semiconductor Library, Vol. 6, Series B, p. 8-58.

Chapter 64

Fig. 64-1: ©Siliconix incorporated, MOSPOWER Design Catalog, 1/83, p. 6-71.
Fig. 64-2: Ferranti Semiconductors, Technical Handbook, Volume 10, Data Converters, 1983. p. 3-12.
Fig. 64-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-144.
Fig. 64-4: Intersil Data Book, 5/83, p. 5-201.
Fig. 64-5: Signetics 555 Timers, 1973, p. 27.
Fig. 64-6: Signetics Analog Data Manual, 1982, p. 6-21.
Fig. 64-7: Signetics Analog Data Manual, 1983, p. 12-36.
Fig. 64-8: Signetics Analog Data Manual, 1983, p. 12-26.
Fig. 64-9: Signetics Analog Data Manual, 1983, p. 12-22.
Fig. 64-10: Electronics Today International, 7/75, p. 39.
Fig. 64-11: Courtesy of Motorola Inc. Circuit Applications for the Triac, AN-466, p. 12.
Fig. 64-13: Electronics Today International, 3/75, p. 67.
Fig. 64-14: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-50.
Fig. 64-15: 73 Magazine, 3/77, p. 152.
Fig. 64-16: Intersil Data Book, 5/83, p. 5-77.
Fig. 64-17: Intersil Data Book, 5/83, p. 5-77.
Fig. 64-18: Intersil Data Book, 5/83, p. 5-77.
Fig. 64-19: Intersil Data Book, 5/83, p. 5-77.
Fig. 64-20: Intersil Data Book, 5/83, p. 5-76.
Fig. 64-21: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-105.
Fig. 64-22: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-15.
Fig. 64-23: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-77.
Fig. 64-24: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-105.
Fig. 64-25: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-105.
Fig. 64-26: Electronics Today International, 6/77, p. 77.
Fig. 64-27: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-15.
Fig. 64-28: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-15.

Fig. 64-29: Signetics Analog Data Manual, 1982, p. 6-14.
Fig. 64-30: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-147.
Fig. 64-31: Electronics Today International, 3/75, p. 67.
Fig. 64-32: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-179.
Fig. 64-33: Signetics Analog Data Manual, 1983, p. 12-28.

Chapter 65

Fig. 65-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 2-8.
Fig. 65-2: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-23.
Fig. 65-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-152.
Fig. 65-4: 101 Electronic Projects, 1975, p. 49.
Fig. 65-5: Electronics Today International, 9/75, p. 64.
Fig. 65-6: Electronics Today International, 3/75, p. 68.
Fig. 65-7: Electronics Today International, 1/75, p. 67.
Fig. 65-8: Reprinted with the permission of National Semiconductor Corp. Voltge Regulator Handbook, p. 10-15.
Fig. 65-9: Electronics Today International, 4/82, p. 29.
Fig. 65-10: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-142.
Fig. 65-11: Signetics Analog Data Manual, 1982, p. 6-25.
Fig. 65-12: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-77.
Fig. 65-13: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-15.
Fig. 65-14: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 1-68.
Fig. 65-15: Reprinted with the permission of National Semiconductor Corp.
Fig. 65-16: Signetics Analog Data Manual, 1982, p. 6-25.
Fig. 65-17: Signetics Analog Data Manual, 1982, p. 6-25.
Fig. 65-18: Electronics Today International, 8/78, p. 91.
Fig. 65-19: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 4-15.
Fig. 65-20: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-147.
Fig. 65-21: Reprinted with the permission of National Semiconductor Corp. CMOS Databook, 1981, p. 6-38.

Chapter 66

Fig. 66-1: No reference.
Fig. 66-2: 73 Magazine.
Fig. 66-3: Electronics Today International, 3/77, p. 71.
Fig. 66-4: Courtesy of Motorola Inc. Circuit Applications for the Triac, AN-466, p. 14.
Fig. 66-5: Electronics Today International, 1/79, p. 95.
Fig. 66-6: Electronics Today International, 8/76, p. 66.
Fig. 66-7: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 2-39.

Chapter 67

Fig. 67-1: Ham Radio, 8/80, p. 18.
Fig. 67-2: Canadian Projects Number 1. p. 86.
Fig. 67-3: Electronics Today International, 5/77, p. 37.
Fig. 67-4: Electronics Today International, 3/81, p.19.
Fig. 67-5: 101 Electronic Projects, 1975, p. 47.
Fig. 67-6: Electronics Today International, 1/76, p. 52.
Fig. 67-7: Electronics Today International, 1/76, p. 51.
Fig. 67-8: Electronics Today International, 11/75, p. 74.
Fig. 67-9: Ham Radio, 2/73, p. 56.
Fig. 67-10: 73 Magazine, 10/83, p. 66.
Fig. 67-11: Electronics Today International, 6/79, p. 103.
Fig. 67-12: Electronics Today International, 1/76, p. 44.
Fig. 67-13: Reprinted from Electronics, 7/76, p. 121. Copyright 1976, McGraw Hill Inc. All rights reserved.

Chapter 68

Fig. 68-1: ©Siliconix incorporated, Analog Switch & IC Product Data Book, 1/82, p. 6-20.
Fig. 68-2: Electronics Today International, 6/79, p. 17.
Fig. 68-3: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B, p. 5-52.
Fig. 68-4: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 445.
Fig. 68-5: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-241.
Fig. 68-6: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-48.
Fig. 68-7: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-24.
Fig. 68-8: Signetics Analog Data Manual, 1982, p. 16-29.
Fig. 68-9: Signetics Analog Data Manual, 1982, p. 16-29.
Fig. 68-10: Teledyne Semiconductor, Databook, p. 8.
Fig. 68-11: © Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 6-20.

Chapter 69

Fig. 69-1: Reprinted from Electronics, 3/75, p. 117. Copyright 1975, McGraw Hill Inc. All rights reserved.
Fig. 69-2: Reprinted from Electronics, 8/78, p. 106. Copyright 1978, McGraw Hill Inc. All rights reserved.
Fig. 69-3: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 2-15.
Fig. 69-4: 49 Easy To Build Projects, TAB Book No. 1337, p. 77.
Fig. 69-5: Electronics Today International, 1/79, p. 97.
Fig. 69-6: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 2-16.

Chapter 70

Fig. 70-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 7-12.
Fig. 70-2: Courtesy of Motorola Inc. Linear Integrated Circuits, p. 6-49.
Fig. 70-3: Ferranti. Technical Handbook Vol. 10, Data Converters, 1983, p. 7-13.
Fig. 70-4: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 4-23.

Chapter 71

Fig. 71-1: Intersil Data Book, 5/83, p. 7-83.
Fig. 71-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-67.
Fig. 71-3: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-37.

Fig. 71-4: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 3-16.
Fig. 71-5: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 13-17.
Fig. 71-6: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-77.
Fig. 71-7: 73 Magazine.
Fig. 71-8: ©Siliconix incorporated, Analog Switch & IC Product Data Book, 1/82, p. 6-18.
Fig. 71-9: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-123.
Fig. 71-10: Ham Radio, 7/76, p. 69.

Chapter 72

Fig. 72-1: 73 Magazine.
Fig. 72-2: CQ, 6/78, p. 32.
Fig. 72-3: Teledyne Semiconductor, Databook, p. 11.
Fig. 72-4: Reprinted from Electronics 4/76, p. 104. Copyright , McGraw Hill Inc. All rights reserved.
Fig. 72-5: Reprinted by permission from the Aug. 1981 issue of Insulation/Circuits magazine. Copyright 1981, Lake Publishing Corporation, Libertyville, Illinois, 60048-9989, USA.
Fig. 72-6: ©Siliconix incorporated, Application Note AN154.
Fig. 72-7: Electronics Today International, 11/78, p. 68.
Fig. 72-8: CQ, 6/78, p.33.

Chapter 73

Fig. 73-1: Courtesy of Motorola Inc. Communications Engineering Bulletin EB-67.
Fig. 73-2: Courtesy of Motorola Inc. Communications Engineering Bulletin EB-63.
Fig. 73-3: Courtesy of Motorola Inc. Application Note AN593, p. 3.
Fig. 73-4: Courtesy of Motorola Inc. Application Note AN-593, p. 6.
Fig. 73-5: Courtesy of Motorola Inc. Communications Engineering Bulletin EB-46.
Fig. 73-6: Microwaves & RF, 1/83, p. 89.
Fig. 73-7: ©Siliconix incorporated, Small Signal FET Design Catalog, 7/83, p. 5-52.
Fig. 73-8: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 7-54.
Fig. 73-9: Wireless World, 11/79, p. 76.
Fig. 73-10: 101 Electronic Projects, 1975, p. 3.
Fig. 73-11: Ham Radio, 10/78, p. 38.
Fig. 73-12: 73 Magazine, 4/83, p. 106.
Fig. 73-13: Ham Radio, 1/74, p. 67.
Fig. 73-14: Courtesy of Motorola Inc. Motorola Semiconductor Library, Vol. 6, Series B, p. 8-59.
Fig. 73-15: © Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 5-36.
Fig. 73-16: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-33.
Fig. 73-17: © Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 5-10.
Fig. 73-18: Reprinted with the permission of National Semiconductor Corp. Application Note 32, p. 9.
Fig. 73-19: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-178.
Fig. 73-20: Signetics Analog Data Manual, 1983, p. 17-13.
Fig. 73-21: Signetics Analog Data Manual, 1983, p. 17-15.
Fig. 73-22: 73 Magazine.
Fig. 73-23: Courtesy of Motorola Inc. Motorola Semiconductor Library, Vol. 6, Series B, p. 8-58.
Fig. 73-24: Courtesy of Motorola Inc. Motorola Semiconductor Library, Vol. 6, Series B, p. 8-58.
Fig. 73-25: © Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 5-10.
Fig. 73-26: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-178.
Fig. 73-27: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-178.
Fig. 73-28: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-178.

Chapter 74

Fig. 74-1: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 8-63.
Fig. 74-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-32.
Fig. 74-3: © Siliconix incorporated. MOSPOWER Design Catalog, 1/83, p. 5-6.
Fig. 74-4: The Giant Book Of Electronics Projects, TAB Book No. 1367.
Fig. 74-5: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 12-14.
Fig. 74-6: Radio-Electronics, 7/83, p. 7.
Fig. 74-7: Radio-Electronics, 7/83, p. 7.
Fig. 74-8: 73 Magazine, 7/77, p. 35.

Chapter 75

Fig. 75-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-126.
Fig. 75-2: Courtesy of Motorola Inc. Communications Engineering Bulletin, EB-46.
Fig. 75-3: Signetics Analog Data Manual, p. 556.
Fig. 75-4: Modern Electronics, 7/78, p. 55.
Fig. 75-5: Electronics Today International, 6/79, p. 43.
Fig. 75-6: Radio-Electronics, 8/69, p. 74.
Fig. 75-7: Signetics 555 Timers, 1973, p. 25.

Chapter 76

Fig. 76-1: The Build-It Book Of Electronic Projects, TAB Book No. 1498, p. 20.
Fig. 76-2: 303 Dynamic Electronic Circuits, TAB Book No. 1060, p. 153.
Fig. 76-3: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-100.
Fig. 76-4: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 225.
Fig. 76-5: '73 Magazine, 9/75, p. 105.
Fig. 76-6: Howard S. Leopold.
Fig. 76-7: Modern Electronics, 3/78, p. 50.
Fig. 76-8: 73 Magazine, 6/83, p. 106.
Fig. 76-9: Modern Electronics, 2/78, p. 50.

Chapter 77

Fig. 77-1: Electronics Today International.
Fig. 77-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-30.
Fig. 77-3: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-31.
Fig. 77-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-18.
Fig. 77-5: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-325.
Fig. 77-6: Reprinted with the permis-

sion of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-152.
Fig. 77-7: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-25.
Fig. 77-8: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-25.
Fig. 77-9: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-25.
Fig. 77-10: Signetics Analog Data Manual, 1982, p. 3-50.
Fig. 77-11: Signetics Analog Data Manual, 1982, p. 3-15.
Fig. 77-12: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-159.

Chapter 78

Fig. 78-1: Electronics Today International, 9/72, p. 86.
Fig. 78-2: Electronics Today International, 1978.
Fig. 78-3: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. 9-76.
Fig. 78-3: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 2-96.
Fig. 78-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-17.

Chapter 79

Fig. 79-1: Supertex Data Book, 1983, p. 5-26.
Fig. 79-2: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. 9-75.
Fig. 79-3: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. 9-76.

Chapter 80

Fig. 80-1: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-40.
Fig. 80-2: Reprinted with the permission of National Semiconductor Corp. COPS Microcontrollers Databook, 1982, p. 9-123.
Fig. 80-3: Reprinted with the permission of National Semiconductor Corp. COPS Microcontrollers Databook, 1982, p. 10-3.
Fig. 80-4: Electronics Today International, 4/78, p. 31.
Fig. 80-5: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-8.
Fig. 80-6: Electronics Today International, 1/79, p. 68.
Fig. 80-7: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-136.
Fig. 80-8: Electronics Today International, 4/78, p. 29.
Fig. 80-9: Electronics Today International, 1/76, p. 49.
Fig. 80-10: Courtesy of Texas Instruments Incorporated. Bulletin No. DL-S 12612, p. 14.
Fig. 80-11: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-9.
Fig. 80-12: Courtesy of Texas Instruments Incorporated. Bulletin No. DL-S 12612, p. 12.

Chapter 81

Fig. 81-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-204.
Fig. 81-2: 73 Magazine, 10/77, p. 115.
Fig. 81-3: Electronics Today International, 7/81, p. 75.
Fig. 81-4: Reprinted with permission from General Electric Semiconductor Department. GE Application Note 200.35, 3/66, p. 14.
Fig. 81-5: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 64.
Fig. 81-6: Electronics Today International, 1975, p. 72.

Chapter 82

Fig. 82-1: Teledyne Semiconductor, Databook, p. 8.
Fig. 82-2: ©Siliconix incorporated. Application Note AN154.
Fig. 82-3: The Complete Handbook of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 335.
Fig. 82-4: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 9-28.
Fig. 82-5: Reprinted from Electronics, 2/77, p. 107. Copyright 19 , McGraw Hill Inc. All rights reserved.
Fig. 82-6: © Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 6-19.
Fig. 82-7: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 2-96.
Fig. 82-8: Electronics Today International, 7/78, p. 16.
Fig. 82-9: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, p. 7-30.
Fig. 82-10: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 13-50.
Fig. 82-11: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, 1979, p. 7-9.
Fig. 82-12: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 145.
Fig. 82-13: Electronics Today International, 7/78, p. 16.
Fig. 82-14: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-31.

Chapter 83

Fig. 83-1: Electronics Today International, 7/81, p. 72.
Fig. 83-2: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 233.
Fig. 83-3: 101 Electronic Projects, 1977, p. 40.

Chapter 84

Fig. 84-1: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-32.
Fig. 84-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-33.
Fig. 84-3: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-28.
Fig. 84-4: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.
Fig. 84-5: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 313.
Fig. 84-6: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 11-10.
Fig. 84-7: Reprinted with permission from General Electric Semiconductor Department. Optoelectronics, Second Edition, p. 141.

Chapter 85

Fig. 85-1: Intersil Data Book, 5/83, p. 7-48.
Fig. 85-2: Reprinted from Electronics, 11/75, p. 120. Copyright 1975, McGraw Hill Inc. All rights reserved.
Fig. 85-3: Courtesy of Motorola Inc.
Fig. 85-4: Mitel Databook, p. 2-17.
Fig. 85-5: Mitel Databook, p. 2-13.
Fig. 85-6: 73 Magazine, 12/83, p. 115.
Fig. 85-7: Ham Radio, 2/77, p. 70.
Fig. 85-8: Ham Radio, 8/77, p. 41.

sion of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-152.
Fig. 77-7: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-25.
Fig. 77-8: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-25.
Fig. 77-9: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-25.
Fig. 77-10: Signetics Analog Data Manual, 1982, p. 3-50.
Fig. 77-11: Signetics Analog Data Manual, 1982, p. 3-15.
Fig. 77-12: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-159.

Chapter 78

Fig. 78-1: Electronics Today International, 9/72, p. 86.
Fig. 78-2: Electronics Today International, 1978.
Fig. 78-3: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. 9-76.
Fig. 78-3: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 2-96.
Fig. 78-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-17.

Chapter 79

Fig. 79-1: Supertex Data Book, 1983, p. 5-26.
Fig. 79-2: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. 9-75.
Fig. 79-3: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. 9-76.

Chapter 80

Fig. 80-1: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 4-40.
Fig. 80-2: Reprinted with the permission of National Semiconductor Corp. COPS Microcontrollers Databook, 1982, p. 9-123.
Fig. 80-3: Reprinted with the permission of National Semiconductor Corp. COPS Microcontrollers Databook, 1982, p. 10-3.
Fig. 80-4: Electronics Today International, 4/78, p. 31.
Fig. 80-5: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-8.
Fig. 80-6: Electronics Today International, 1/79, p. 68.
Fig. 80-7: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-136.
Fig. 80-8: Electronics Today International, 4/78, p. 29.
Fig. 80-9: Electronics Today International, 1/76, p. 49.
Fig. 80-10: Courtesy of Texas Instruments Incorporated. Bulletin No. DL-S 12612, p. 14.
Fig. 80-11: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 5-9.
Fig. 80-12: Courtesy of Texas Instruments Incorporated. Bulletin No. DL-S 12612, p. 12.

Chapter 81

Fig. 81-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-204.
Fig. 81-2: 73 Magazine, 10/77, p. 115.
Fig. 81-3: Electronics Today International, 7/81, p. 75.
Fig. 81-4: Reprinted with permission from General Electric Semiconductor Department. GE Application Note 200.35, 3/66, p. 14.
Fig. 81-5: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 64.
Fig. 81-6: Electronics Today International, 1975, p. 72.

Chapter 82

Fig. 82-1: Teledyne Semiconductor, Databook, p. 8.
Fig. 82-2: ©Siliconix incorporated. Application Note AN154.
Fig. 82-3: The Complete Handbook of Amplifiers, Oscillators & Multivibrators, TAB Book No. 1230, p. 335.
Fig. 82-4: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 9-28.
Fig. 82-5: Reprinted from Electronics, 2/77, p. 107. Copyright 19 , McGraw Hill Inc. All rights reserved.
Fig. 82-6: © Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 6-19.
Fig. 82-7: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 2-96.
Fig. 82-8: Electronics Today International, 7/78, p. 16.
Fig. 82-9: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, p. 7-30.
Fig. 82-10: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 13-50.
Fig. 82-11: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, 1979, p. 7-9.
Fig. 82-12: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 145.
Fig. 82-13: Electronics Today International, 7/78, p. 16.
Fig. 82-14: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-31.

Chapter 83

Fig. 83-1: Electronics Today International, 7/81, p. 72.
Fig. 83-2: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 233.
Fig. 83-3: 101 Electronic Projects, 1977, p. 40.

Chapter 84

Fig. 84-1: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-32.
Fig. 84-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-33.
Fig. 84-3: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-28.
Fig. 84-4: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.
Fig. 84-5: Reprinted with permission from General Electric Semiconductor Department. General Electric SCR Manual, Sixth Edition, 1979, p. 313.
Fig. 84-6: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 11-10.
Fig. 84-7: Reprinted with permission from General Electric Semiconductor Department. Optoelectronics, Second Edition, p. 141.

Chapter 85

Fig. 85-1: Intersil Data Book, 5/83, p. 7-48.
Fig. 85-2: Reprinted from Electronics, 11/75, p. 120. Copyright 1975, McGraw Hill Inc. All rights reserved.
Fig. 85-3: Courtesy of Motorola Inc.
Fig. 85-4: Mitel Databook, p. 2-17.
Fig. 85-5: Mitel Databook, p. 2-13.
Fig. 85-6: 73 Magazine, 12/83, p. 115.
Fig. 85-7: Ham Radio, 2/77, p. 70.
Fig. 85-8: Ham Radio, 8/77, p. 41.

Fig. 85-9: Ham Radio, 1/84, p. 94.
Fig. 85-10: Reprinted with permission from General Electric Semiconductor Department. Optoelectronics, Second Edition, p. 119.
Fig. 85-11: Signetics Analog Data Manual, 1982, p. 16-27.
Fig. 85-12: Modern Electronics, 7/78, p. 56.
Fig. 85-13: The Build-It Book Of Electronic Projects, TAB Book No. 1498, p. 3.
Fig. 85-14: Reprinted with the permission of National Semiconductor Corp. COPS Microcontrollers Databook, 1982, p. 9-118.
Fig. 85-15: 73 Magazine, 1/84, p. 115.
Fig. 85-16: Intersil Data Book, 5/83, p. 7-47.
Fig. 85-17: Reprinted with permission from General Electric Semiconductor Department Optoelectronics, Second Edition, p. 119.
Fig. 85-18: Ham Radio, 1/84, p. 93.
Fig. 85-19: Ham Radio, 1/84, p. 91.
Fig. 85-20: 73 Magazine, 4/83.
Fig. 85-21: 73 Magazine, 9/82, p. 92.

Chapter 86

Fig. 86-1: Radio-Electronics, 7/81, p. 73.
Fig. 86-2: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-87.
Fig. 86-3: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 12-17.
Fig. 86-4: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-162.
Fig. 86-5: Courtesy of Motorola Inc. Circuit Applications for the Triac (AN-466), p. 9.
Fig. 86-6: Courtesy of Motorola Inc. Circuit Applications for the Triac, AN-466, p. 13.
Fig. 86-7: Intersil Data Book, 5/83, p. 5-68.
Fig. 86-8: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. LB36-2.
Fig. 86-9: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-29.
Fig. 86-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-6.
Fig. 86-11: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-29.
Fig. 86-12: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 7-33.

Chapter 87

Fig. 87-1: Electronics Today International, 4/81, p. 86.
Fig. 87-2: Electronics Today International, 12/78, p. 32.
Fig. 87-3: Signetics Analog Data Manual, 1983, p. 10-65.
Fig. 87-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-147.
Fig. 87-5: Teledyne Semiconductor, Databook, p. 12.
Fig. 87-6: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 10-16.
Fig. 87-7: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 12-9.
Fig. 87-8: Signetics Analog Data Manual, 1982, p. 3-78.
Fig. 87-9: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 12-7.
Fig. 87-10: Radio-Electronics, 3/80, p. 60.
Fig. 87-11: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 12-10.
Fig. 87-12: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-162.
Fig. 87-13: Intersil Data Book, 5/83, p. 5-71.
Fig. 87-14: Intersil Data Book, 5/83, p. 5-71.
Fig. 87-15: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-42.
Fig. 87-16: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-108.
Fig. 87-17: Reprinted with the permission of National Semiconductor Corp. CMOS Databook, 1981, p. 6-7.
Fig. 87-18: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-31.
Fig. 87-19: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-31.
Fig. 87-20: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-29.
Fig. 87-21: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-160.
Fig. 87-22: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-162.
Fig. 87-23: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-107.
Fig. 87-24: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 2-46.
Fig. 87-25: Electronics Today International, 10/78, p. 101.
Fig. 87-26: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 2-46.
Fig. 87-27: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-29.
Fig. 87-28: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-160.
Fig. 87-29: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-31.
Fig. 87-30: Teledyne Semiconductor, Databook, p. 11.
Fig. 87-31: Teledyne Semiconductor, Databook, p. 11.
Fig. 87-32: Intersil Data Book, 5/83, p. 5-70.
Fig. 87-33: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-29.

Chapter 88

Fig. 88-1: Western Digital, Components Handbook, 1983, p. 579.
Fig. 88-2: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 289.
Fig. 88-3: Signetics Analog Data Manual, 1983, p. 15-11.
Fig. 88-4: Courtesy of Motorola Inc. Application Note AN-294, p. 6.
Fig. 88-5: Reprinted with permission from General Electric Semiconductor Department. Application Note 201.11.
Fig. 88-6: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 1183.
Fig. 88-7: Signetics 555 Timers, 1973, p. 19.
Fig. 88-8: RCA Corporation, Linear Integrated Circuits And MOS/FETS, p. 437.
Fig. 88-9: Reprinted with permission from General Electric Semiconductor

Department. GE Semiconductor Data Handbook, Second Edition, p. 412.
Fig. 88-10: 73 Magazine, 8/75, p. 140.
Fig. 88-11: Western Digital, Components Handbook, 1983, p. 581.
Fig. 88-12: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Second Edition, p. 727.
Fig. 88-13: Electronics Today International, 3/82, p. 67.
Fig. 88-14: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-17.
Fig. 88-15: Electronics Today International, 1/76, p. 52.
Fig. 88-16: Modern Electronics, 2/78, p. 49.
Fig. 88-17: Signetics 555 Timers, 1973, p. 26.
Fig. 88-18: Signetics 555 Timers, 1973, p. 20.

Chapter 89

Fig. 89-1: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-170.
Fig. 89-2: Signetics Analog Data Manual, 1982, p. 3-89.
Fig. 89-3: Electronics Today International, 10/77, p. 34.
Fig. 89-4: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 130.
Fig. 89-5: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-63.
Fig. 89-6: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-53.
Fig. 89-7: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-49.
Fig. 89-8: Electronics Today International, 6/79, p. 105.
Fig. 89-9: Electronics Today International, 6/82, p. 66.
Fig. 89-10: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 130.
Fig. 89-11: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-27.
Fig. 89-12: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-48.
Fig. 89-13: Electronics Today International.

Chapter 90

Fig. 90-1: Radio-Electronics, 12/81, p. 52.
Fig. 90-2: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-108.
Fig. 90-3: 73 Magazine, 6/77, p. 49.
Fig. 90-4: CQ, 6/83, p. 46.
Fig. 90-5: 73 Magazine, 8/83, p. 100.

Chapter 91

Fig. 91-1: Electronics Today International, 6/78, p. 29.
Fig. 91-2: 73 Magazine, 2/83, p. 90.
Fig. 91-3: Radio-Electronics, 3/80, p. 60.
Fig. 91-4: Radio-Electronics, 8/83, p. 96.
Fig. 91-5: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-11.

Chapter 92

Fig. 92-1: Courtesy of Motorola Inc. Application Note AN-545A, p. 7.
Fig. 92-2: Courtesy of Motorola Inc. Application Note AN-545A, p. 12.
Fig. 92-3: Plessey Semiconductors, Linear IC Handbook, 5/82, p. 129.
Fig. 92-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-50.
Fig. 92-5: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-73.
Fig. 92-6: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 5-51.
Fig. 92-7: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 7-26.
Fig. 92-8: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-31.
Fig. 92-9: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-30.
Fig. 92-10: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B.
Fig. 92-11: Harris Semiconductor, Linear & Data Acquisition Products, 1977, p. 2-46.

Chapter 93

Fig. 93-1: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-59.
Fig. 93-2: Reprinted with the permission of National Semiconductor Corp. Voltage Regulator Handbook, p. 10-47.
Fig. 93-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-23.
Fig. 93-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 7-11.
Fig. 93-5: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-158.
Fig. 93-6: Signetics Analog Data Manual, 1982, p. 3-38.
Fig. 93-7: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 13-50.
Fig. 93-8: Courtesy of Motorola Inc., Linear Integrated Circuits, 1979, p. 3-42.
Fig. 93-9: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-25.
Fig. 93-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-142.
Fig. 93-11: Precision Monolithics Incorporated, 1981 Full Line Catalog,p. 10-18.
Fig. 93-12: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 10-15.
Fig. 93-13: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-16.
Fig. 93-14: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 10-8.
Fig. 93-15: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 14-52.
Fig. 93-16: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-158.
Fig. 93-17: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 14-44.
Fig. 93-18: Electronics Today International, 8/78, p. 91.
Fig. 93-19: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 14-41.
Fig. 93-20: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-78.
Fig. 93-21: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 14-53.
Fig. 93-22: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 14-53.
Fig. 93-23: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Data-

book, 1980, p. 14-51.

Chapter 94

Fig. 94-1: Intersil Data Book, 5/83, p. 5-238.
Fig. 94-2: Reprinted with the permission of National Semiconductor Corp. Data Databook, 1982, p. 5-9.
Fig. 94-3: Electronics Today International, 12/78, p. 20.
Fig. 94-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-17.
Fig. 94-5: Electronics Today International, 7/72, p. 84.
Fig. 94-6: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-13.
Fig. 94-7: Signetics Analog Data Manual, 1982, p. 8-14.
Fig. 94-8: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-179.
Fig. 94-9: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-238.

Chapter 95

Fig. 95-1: Teledyne Semiconductor, Publication DG-114-87, p. 3.
Fig. 95-2: ©Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 1-25.
Fig. 95-3: Courtesy of Fairchild Camera & Instrument Ctrporation. Linear Databook, 1982, p. 7-7.
Fig. 95-4: Reprinted with the permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 12-19.
Fig. 95-5: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. D-7.
Fig. 95-6: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 12-20.

Chapter 96

Fig. 96-1: Reprinted with the permission of National Semiconductor Corp. National Semiconductor CMOS Databook, 1981, p. 3-50.
Fig. 96-2: Precision Monolithics Incorporated, 1981, Full Line Catalog, p. 16-138.
Fig. 96-3: Teledyne Semiconductor, Databook, p. 11.
Fig. 96-4: ©Siliconix incorporated, Analog Switch & IC Product Data Book, 1/82, p. 7-21.
Fig. 96-5: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-141.
Fig. 96-6: Reprinted with the permission of National Semiconductor Corp. Application Note 32, p. 2.
Fig. 96-7: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 9-204.
Fig. 96-8: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-103.
Fig. 96-9: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 17-54.
Fig. 96-10: Electronics Today International, 7/72, p. 83.
Fig. 96-11: Signetics Analog Data Manual, 1982, p. 3-50.
Fig. 96-12: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 1-7.
Fig. 96-13: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-27.

Chapter 97

Fig. 97-1: Reprinted from Electronics, 7/83, p. 135. Copyright 1983, McGraw Hill Inc. All rights reserved.
Fig. 97-2: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 8-33.
Fig. 97-3: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-173.
Fig. 97-4: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 145.
Fig. 97-5: Intersil Data Book, 5/83, p. 5-238.
Fig. 97-6: Harris Semiconductor, Linear & Data Acquisition Products, p. 2-58.
Fig. 97-7: 73 Magazine, 8/78, p.132.
Fig. 97-8: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-241.
Fig. 97-9: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 577.
Fig. 97-10: Reprinted with permission from General Electric Semiconductor Department. GE Semiconductor Data Handbook, Third Edition, p. 1183.
Fig. 97-11: Intersil Data Book, 5/83, p. 5-238.
Fig. 97-12: Courtesy of Motorola Inc. Linear Interface Integrated Circuits, 1979, p. 5-119.
Fig. 97-13: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-81.
Fig. 97-14: Harris Semiconductor Linear – Data Acquisition Products, p. 2-46.
Fig. 97-15: Intersil Data Book 5/83, p. 4-93.
Fig. 97-16: Signetics Analog Data Manual, 1982, p. 16-29.
Fig. 97-17: Signetics Analog Data Manual, 1982, p. 16-29.
Fig. 97-18: Signetics Analog Data Manual, 1977, p. 264.
Fig. 97-19: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-25.
Fig. 97-20: Signetics Analog Data Manual, 1982, p. 16-29.

Chapter 98

Fig. 98-1: Electronics Today International, 8/78, p. 69.
Fig. 98-2: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-32.
Fig. 98-3: ©Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 6-18.
Fig. 98-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-123.
Fig. 98-5: Courtesy of Texas Instruments Incorporated. Linear Control Circuis Data Book, Second Edition, p. 205.
Fig. 98-6: ©Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 6-14.

Index

Numbers preceded by an "I-," "II," and "III" are from *Encyclopedia of Electronic Circuits Vol. I., Vol. II, and Vol. III* respectively.

B

C

D

E

F

G

H

I

J

K

M

N

O

P

photo conductive detector amplifier, four quadrant, I-359
photo memory switch for ac power control, I-363
photo stop action, I-481

Q

R

S

T

U

V

W

X

Y

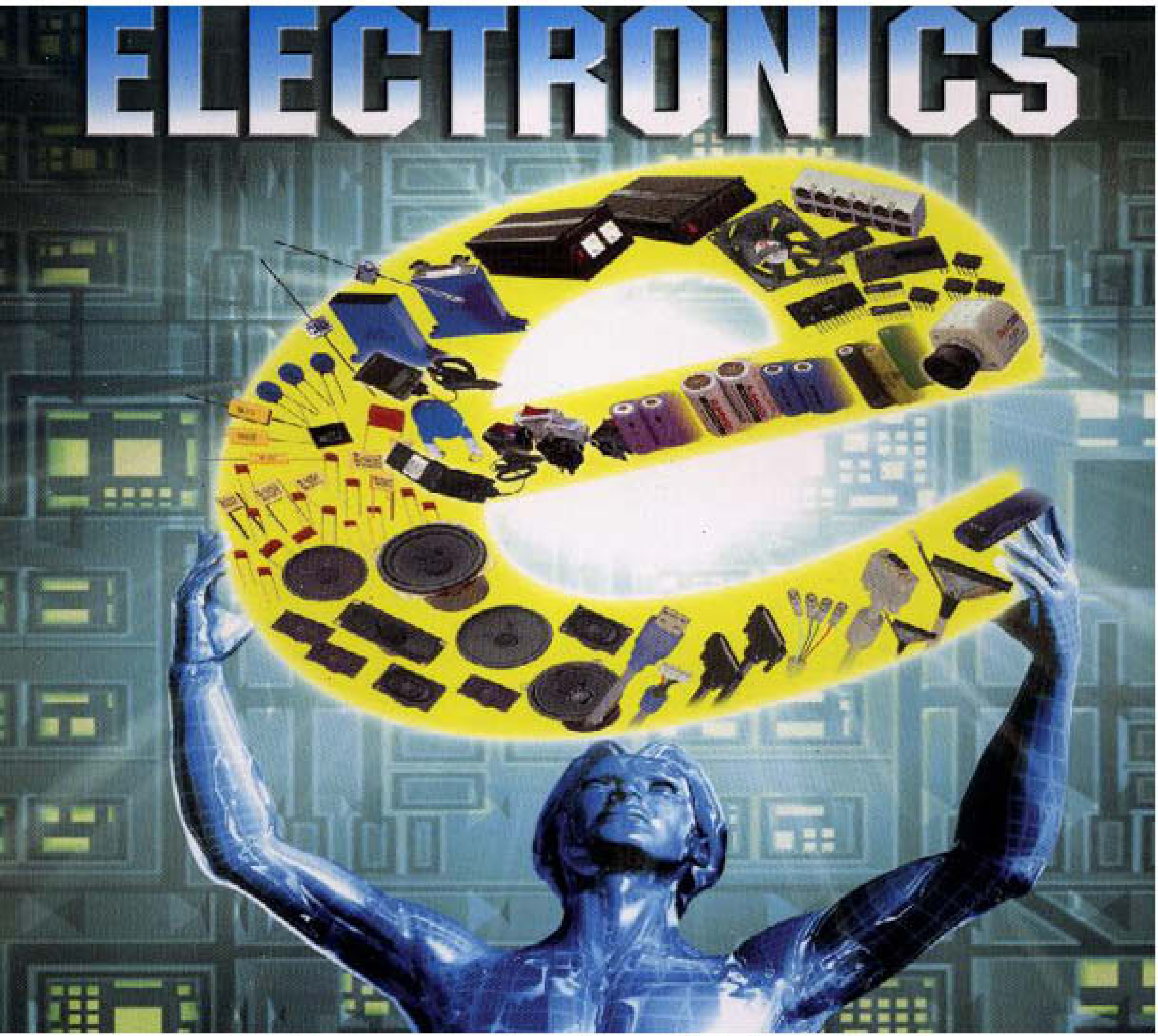
Z

3138

Encyclopedia of ELECTRONIC CIRCUITS

Volume 2

Rudolf F. Graf


ELECTRONICS

CONTENTS

Introduction

Encyclopedia of Electronic Circuits—Volume 2, a companion to Volume 1 published in 1985, contains well over 1400 not-previously covered circuits organized into 108 chapters. For each reference, circuits are listed at the beginning of each chapter. The extensive index further enhances the usefulness of this new work. The browser, as well as the serious researcher looking for a very specific circuit, will be richly rewarded by the context of this volume. A brief explanatory text accompanies almost every entry. The original source for each item is also given so that the reader requiring additional data will know where to find it.

I am most grateful to William Sheets for his many and varied contributions to this book, and to Mrs. Stella Dillon for her fine work at the word processor. These friends and associates of long standing have my sincere thanks for contributing to the successful completion of this book.

1

Alarm and Security Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AUTO BURGLAR ALARM

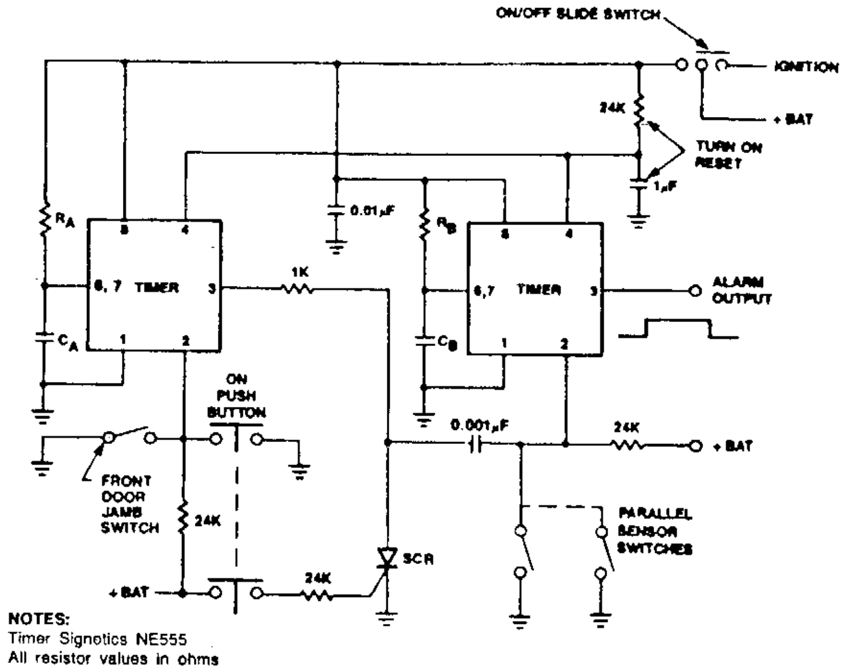


Fig. 1-1

SIGNETICS

Circuit Notes

Timer A produces a safeguard delay, allowing the driver to disarm the alarm and eliminating a vulnerable outside control switch. The SCR prevents timer A from triggering timer B, unless timer B is triggered by strategically-located sensor switches.

MULTIPLE ALARM CIRCUIT

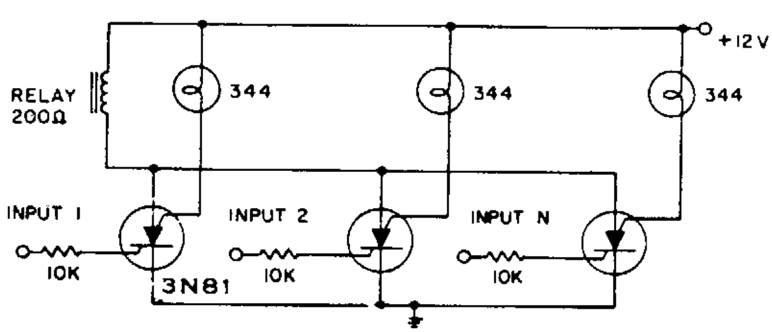


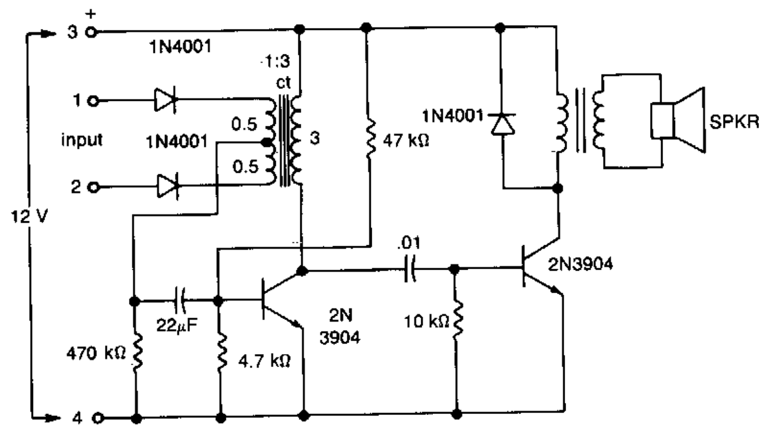
Fig. 1-2

GE

Circuit Notes

Any of several inputs pulls in the common alarm relay with lamps giving visual indication of triggering input. Low resistance lamps decrease input sensitivity.

DIFFERENTIAL VOLTAGE OR CURRENT ALARM



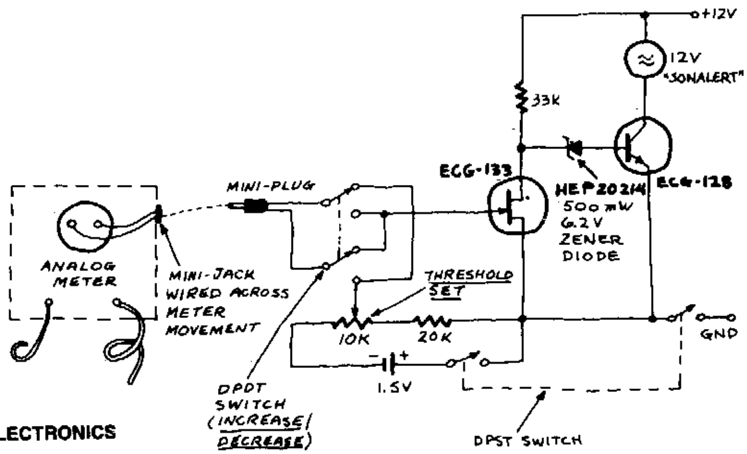
WILLIAM SHEETS

Fig. 1-3

Circuit Notes

The input may be dc or low frequency ac. The output is a distinctive series of audio beeps or a continuous tone, and occurs only when a selected polarity unbalance is present at the input.

TROUBLE TONE ALERT



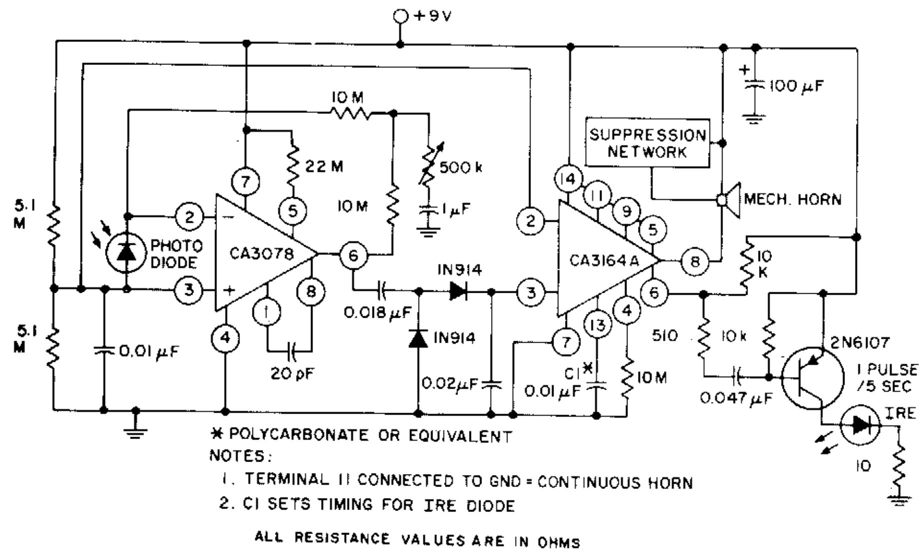
RADIO ELECTRONICS

Fig. 1-4

Circuit Notes

The Trouble Tone Alert is intended for use with analog meters—just wire a "mini" earphone jack directly across the meter movement, plug it in, and you're all set. This device reacts the to the meter-movement driving voltage. It will respond to a change in ac or dc voltage, current, or in resistance. The circuit will respond to an increase or decrease selected by the DPDT switch and is adjusted with the threshold control until the tone from the Sonalert just disappears (with the meter in the circuit being tested, of course).

PHOTOELECTRIC ALARM SYSTEM



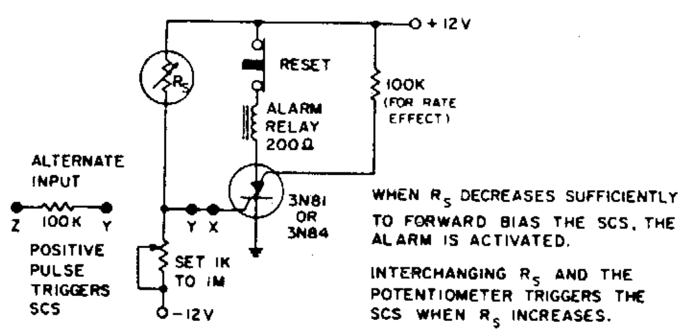
GE/RCA

Fig. 1-5

Circuit Notes

The CA3164A BiMOS detector alarm system and the CA3078 micropower op amp with a photodiode are used as an automatic switch for turning on a night light or sounding a mechanical horn.

ALARM CIRCUIT



GE

Fig. 1-6

Circuit Notes

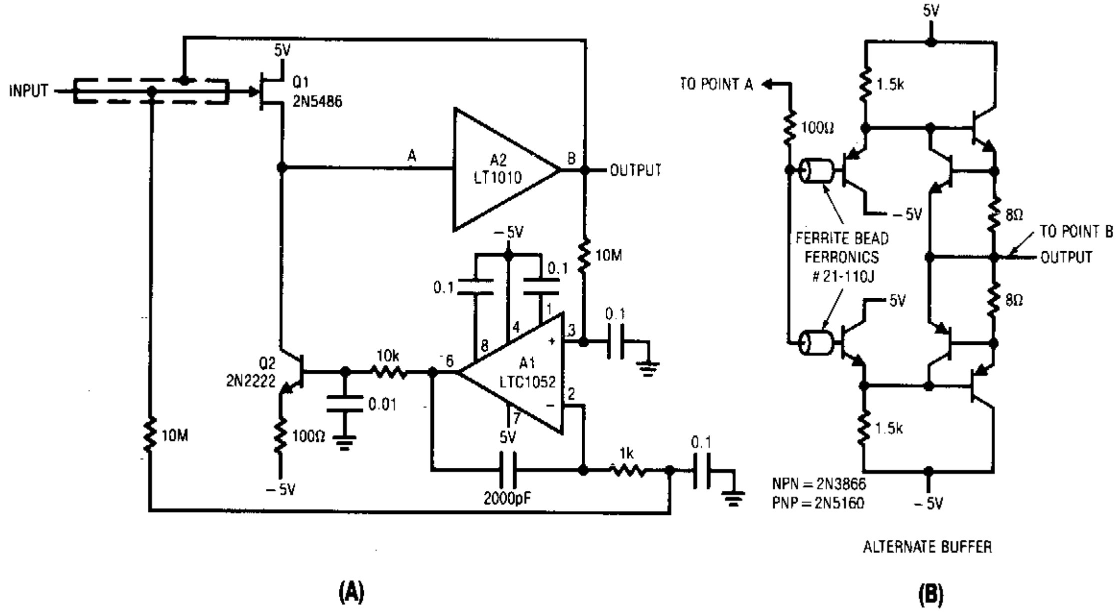
Temperature, light, or radiation sensitive resistors up to 1 megohm readily trigger the alarm when they drop below the value of the preset potentiometer. Alternately, 0.75 V at the input to the 100 kΩ triggers the alarm. Connecting SCS between ground and −12 V permits triggering on negative input to G_A.

2

Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

STABLE UNITY GAIN BUFFER WITH GOOD SPEED AND HIGH INPUT IMPEDANCE



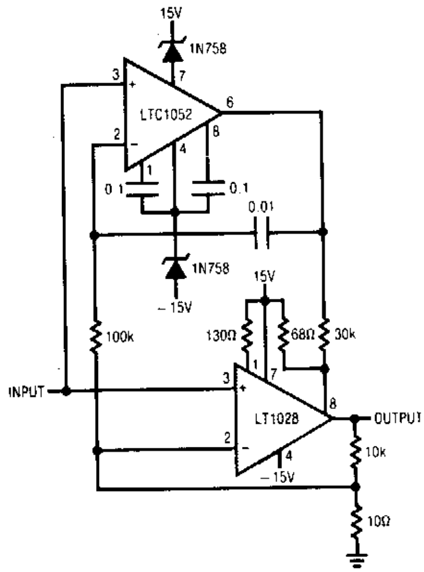
LINEAR TECHNOLOGY CORPORATION

Fig. 2-1

Circuit Notes

Q1 and Q2 constitute a simple, high speed FET input buffer. Q1 functions as a source follower, with the Q2 current source load setting the drain-source channel current. Normally, this open loop configuration would be quite drifty because there is no dc feedback. The LTC1052 contributes this function to stabilize the circuit by comparing the filtered circuit output to a similarly filtered version of the input signal. The amplified difference between these signals is used to set Q2's bias and hence Q1's channel current. This forces Q1's V_{GS} to whatever voltage is required to match the circuit's input and output potentials. The 2000 pF capacitor at A1 provides stable loop compensation. The RC network in A1's output prevents it from seeing high speed edges coupled through Q2's collector-base junction. A2's output is also fed back to the shield around Q1's gate lead, bootstrapping the circuit's effective input capacitance down to less than 1 pF. For very fast requirements, the alternate discrete component buffer shown will be useful. Although its output is current limited at 75 mA, the GHz range transistors employed provide exceptionally wide bandwidth, fast slewing and very little delay.

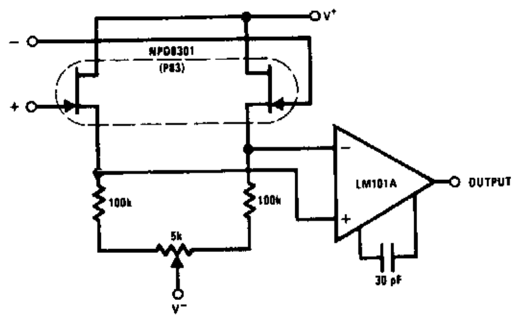
CHOPPER STABILIZED AMPLIFIER



LINEAR TECHNOLOGY CORP.

Fig. 2-2

FET INPUT AMPLIFIER



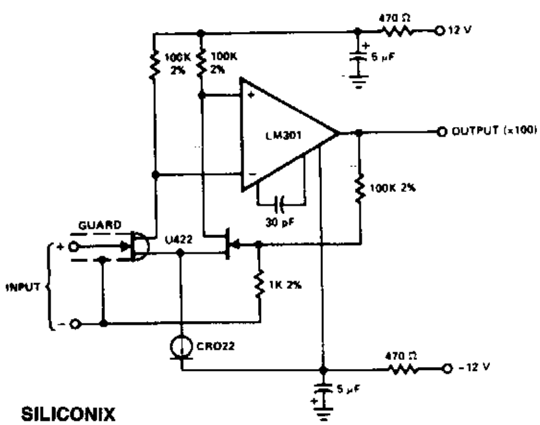
NATIONAL SEMICONDUCTOR CORP.

Circuit Notes

The NPD8301 monolithic-dual provides an ideal low offset, low drift buffer function for the LM101A op amp. The excellent matching characteristics of the NPD8301 track well over its bias current range, thus improving common-mode rejection.

Fig. 2-4

ULTRA-LOW-LEAKAGE PREAMPLIFIER



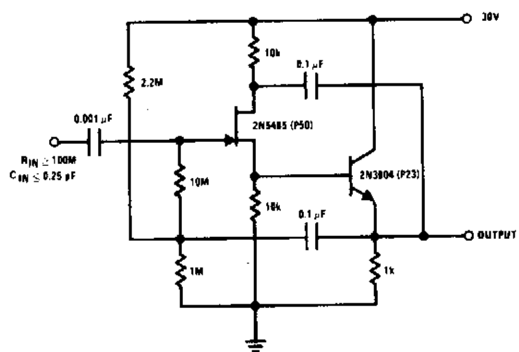
SILICONIX

Circuit Notes

The circuit has an input leakage of only 2 pA typical at 75°C and would be usable with 1 M ohm input resistance.

Fig. 2-3

ULTRA-HIGH Z_{in} AC UNITY GAIN AMPLIFIER



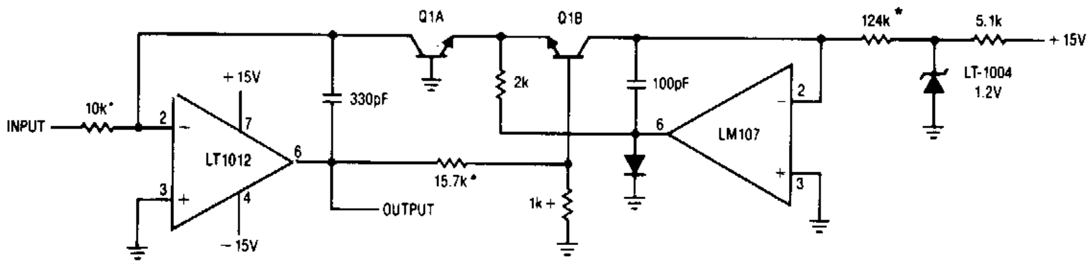
NATIONAL SEMICONDUCTOR CORP.

Circuit Notes

Nothing is left to chance in reducing input capacitance. The 2N5485, which has low capacitance in the first place, is operated as a source follower with bootstrapped gate bias resistor and drain.

Fig. 2-5

LOGARITHMIC AMPLIFIER



+ = TEL. LABS. TYPE Q81
* = 1% FILM RESISTOR
Q1 = 2N2979

Low bias current and offset voltage of the LT1012 allow 4½ decades of voltage input logging.

LINEAR TECHNOLOGY CORP.

Fig. 2-6

COMPOSITE AMPLIFIER

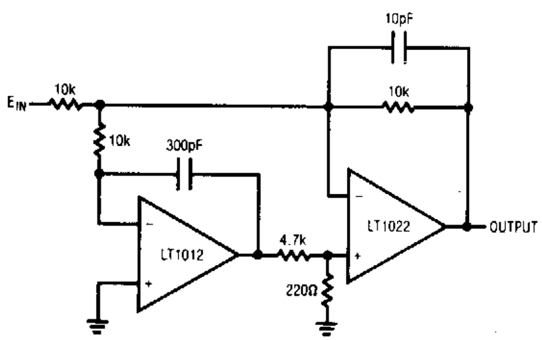


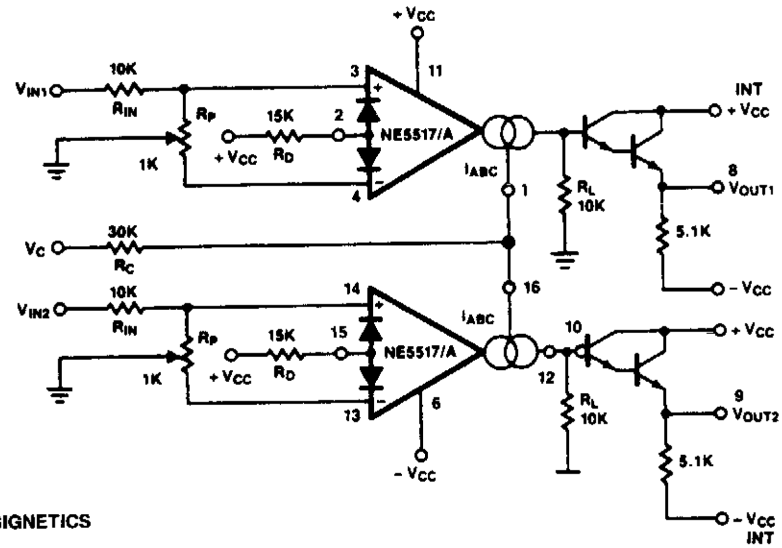
Fig. 2-7

LINEAR TECHNOLOGY CORPORATION

Circuit Notes

The circuit is made up of an LT1012 low drift device, and an LT1022 high speed amplifier. The overall circuit is a unity gain inverter, with the summing node located at the junction of three 10 k resistors. The LT1012 monitors this summing node, compares it to ground, and drives the LT1022's positive input, completing a dc stabilizing loop around the LT1022. The 10 k - 300 pF time constant at the LT1012 limits its response to low frequency signals. The LT1022 handles high frequency inputs while the LT1012 stabilizes the dc operating point. The 4.7 k - 220 ohm divider at the LT1022 prevents excessive input overdrive during start-up. This circuit combines the LT1012's 35 μV offset and 1.5 V/°C drift with the LT1022's 23 V/μs slew rate and 300 kHz full power bandwidth. Bias current, dominated by the LT1012, is about 100 pA.

STEREO AMPLIFIER WITH GAIN CONTROL



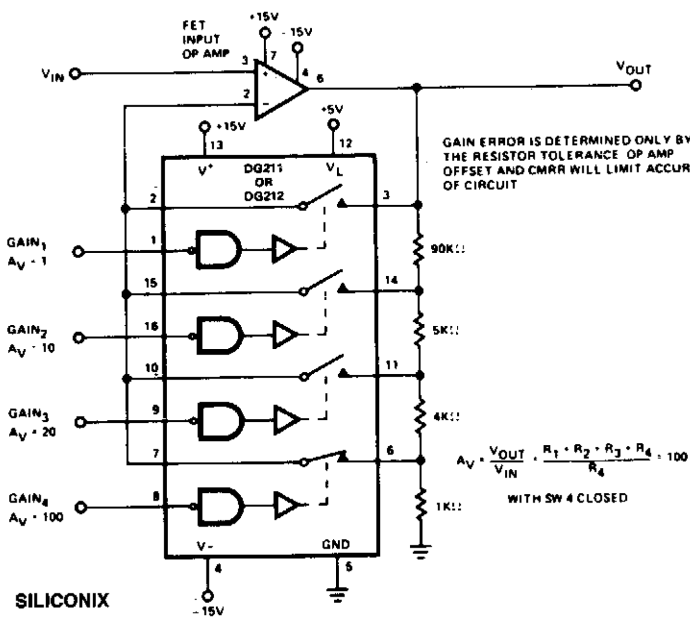
SIGNETICS

Fig. 2-8

Circuit Notes

Excellent tracking of typical 0.3 dB is easy to achieve. With the potentiometer, R_p, the offset can be adjusted. For ac-coupled amplifiers, the potentiometer may be replaced with two 5.1 k ohm resistors.

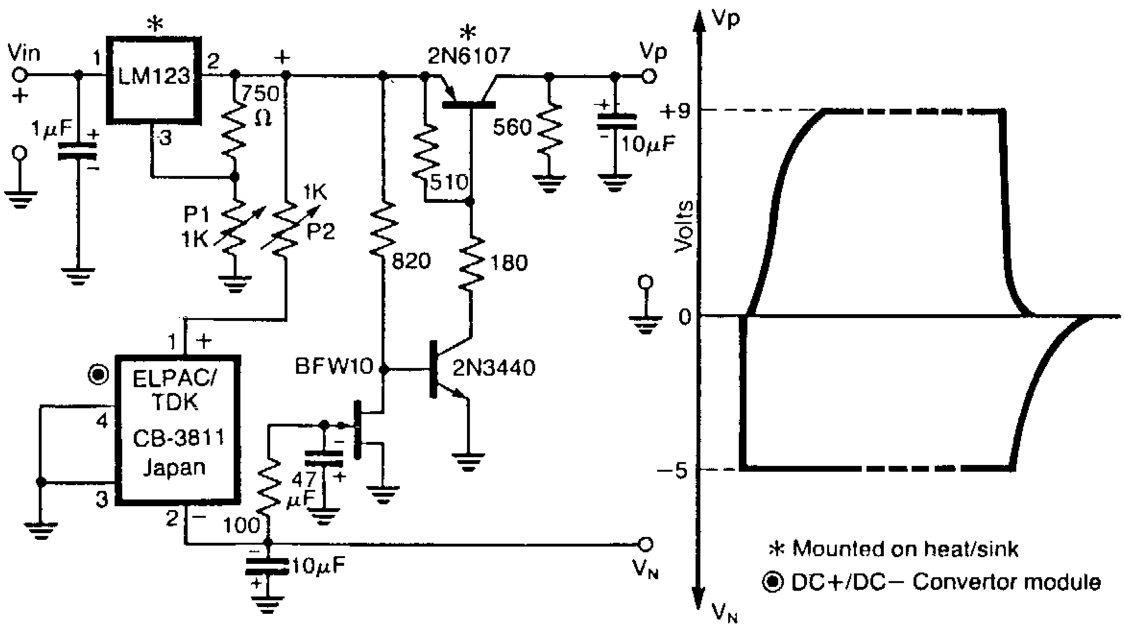
PRECISION-WEIGHTED RESISTOR PROGRAMMABLE-GAIN AMPLIFIER



SILICONIX

Fig. 2-9

POWER GaAsFET AMPLIFIER WITH SINGLE SUPPLY



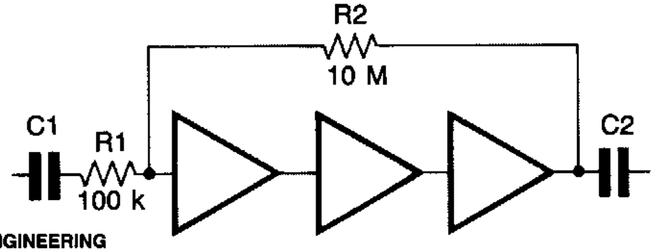
ELECTRIC ENGINEERING

Fig. 2-10

Circuit Notes

The dual regulator circuit operates from a positive supply, which when switched ON powers the gate first, and when switched OFF shuts off the drain first as shown in the figure. This circuit incorporates the LM123, a three terminal positive regulator and a dc+ to dc− converter, the outputs of which power the drains and gates of the power GaAsFETs in a power amplifier relay. The output of the three terminal regulator drives a dc+ to dc− converter whose output biases an N-channel JFET suitably so as to pull the base of the series pass transistor 2N6107 to a level to turn it on. The circuit will turn off the drain supply whenever the negative potential on the Gate fails.

LINEAR AMPLIFIERS FROM CMOS INVERTERS



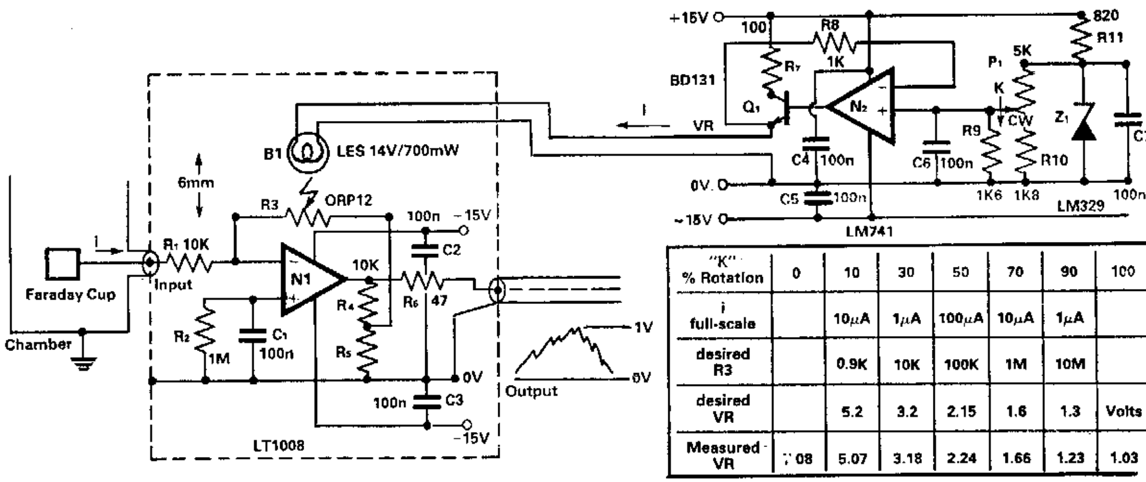
ELECTRONIC ENGINEERING

Fig. 2-11

Circuit Notes

CMOS inverters can be used as linear amplifiers if negative feedback is applied. Best linearity is obtained with feedback applied around three inverters which gives almost perfect linearity up to an output swing of 5 V p-p with a 10 V supply rail. The gain is set by the ratio of R1 and R2 and the values shown are typical for a gain of 100. The high frequency response with the values shown is almost flat to 20 kHz. The frequency response is determined by C1 and C2. This circuit is not suitable for low level signals because the signal-to-noise ratio is only approx. 50 dB with 5 V p-p output with the values shown.

CURRENT-COLLECTOR HEAD-AMPLIFIER



"K" % Rotation	0	10	30	50	70	90	100
i full-scale		10μA	1μA	100μA	10μA	1μA	
desired R3		0.9K	10K	100K	1M	10M	
desired VR		5.2	3.2	2.15	1.6	1.3	Volts
Measured VR	[illegible]08	5.07	3.18	2.24	1.66	1.23	1.03

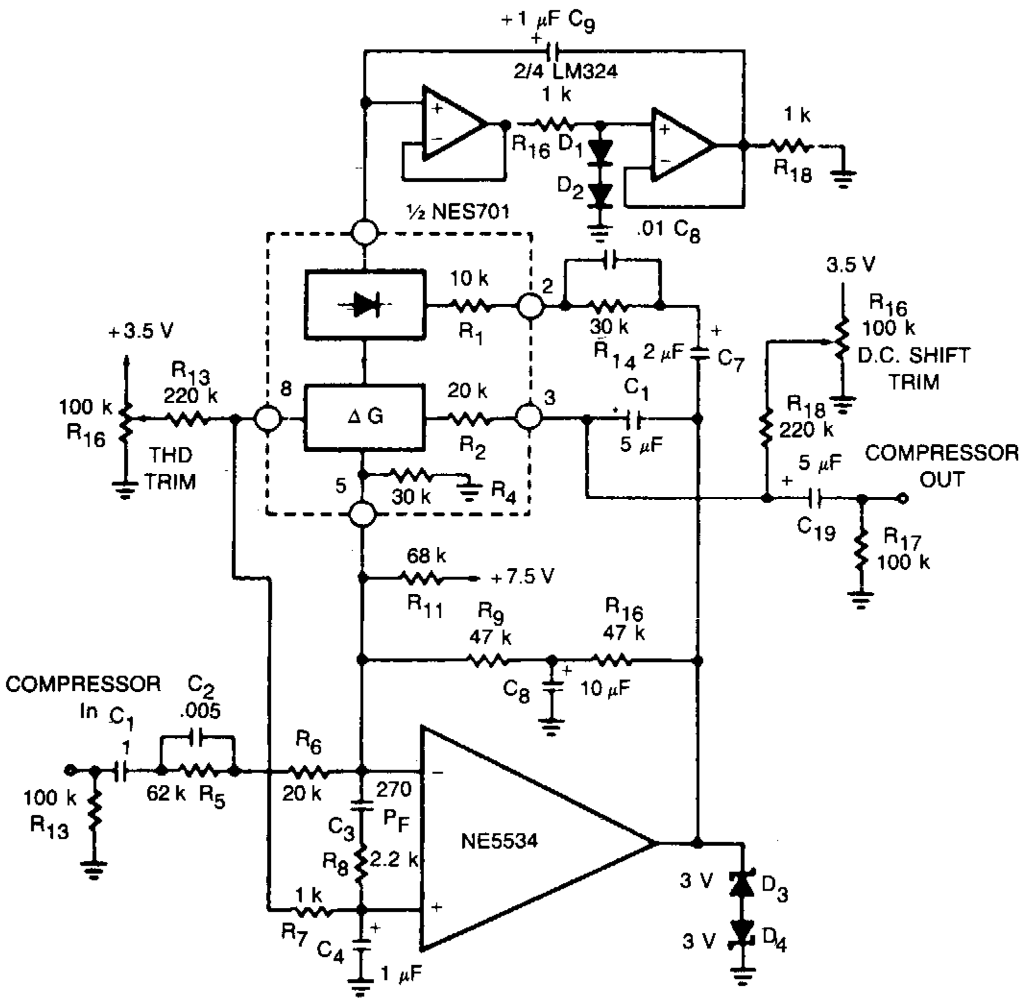
ELECTRONIC ENGINEERING

Fig. 2-12

Circuit Notes

To amplify small current signals such as from an electron-collector inside a vacuum chamber, it is convenient for reasons of noise and bandwidth to have a "head-amplifier" attached to the chamber. The op-amp N_1 is a precision bipolar device with extremely low bias current and offset voltage (1) as well as low noise, which allows the 100:1 feedback attenuator to be employed. The resistance of R_3 can be varied from above 10 M to below 1 k, and so the nominal 0 to 1 V-peak output signal corresponds to input current ranges of 1 nA to 10 μA.

HI-FI COMPANDER



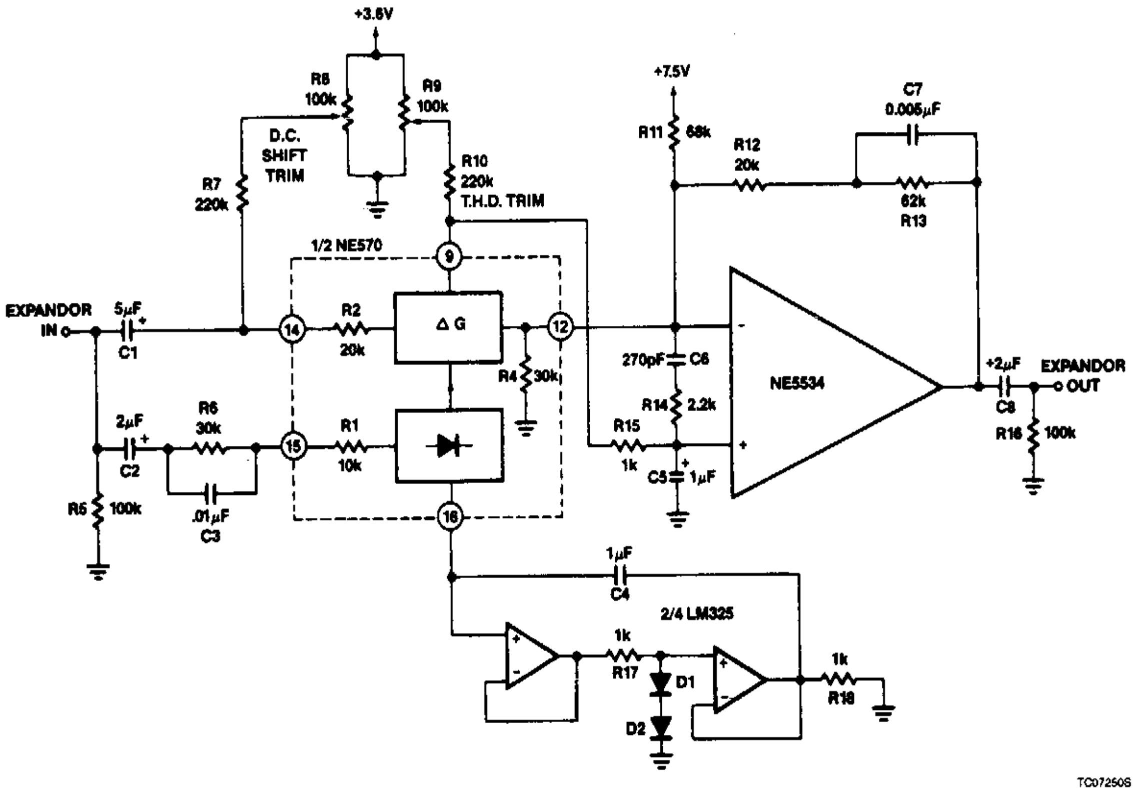
SIGNETICS

HI-FI Compressor With Pre-emphasis

Fig. 2-13(A)

Circuit Notes

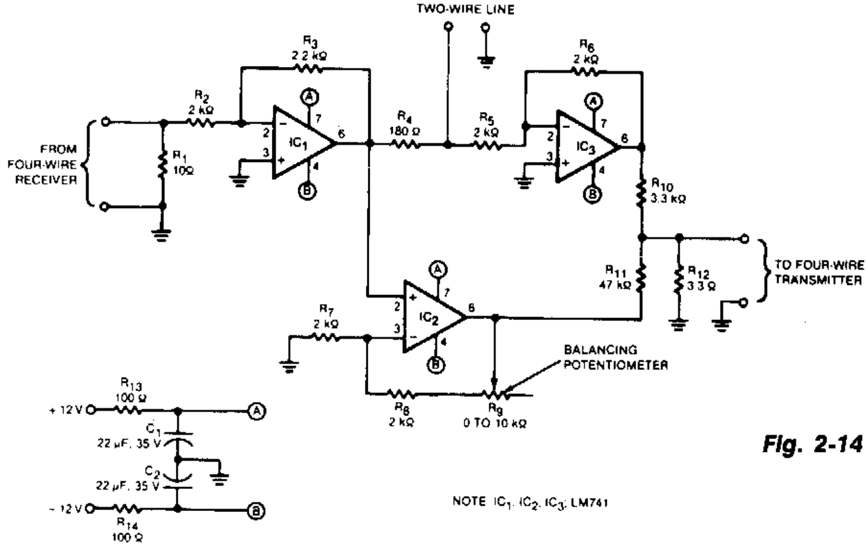
This circuit for a high fidelity compressor uses an external op amp, and has a high gain and wide bandwidth. An input compensation network is required for stability. The rectifier capacitor (C_9) is not grounded, but is tied to the output of an op amp circuit. When a compressor is operating at high gain, (small input signal), and is suddenly hit with a signal, it will overload until



Hi-Fi Expandor With De-emphasis *Fig. 2-13(B)*

it can reduce its gain. The time it takes for the compressor to recover from overload is determined by the rectifier capacitor C_9. The expandor to complement the compressor is shown in Fig. 2-13B. Here an external op amp is used for high slew rate. Both the compressor and expandor have unity gain levels of 0 dB. Trim networks are shown for distortion (THD) and dc shift. The distortion trim should be done first, with an input of 0 dB at 10 kHz. The dc shift should be adjusted for minimum envelope bounce with tone bursts.

TWO-WIRE TO FOUR-WIRE AUDIO CONVERTER



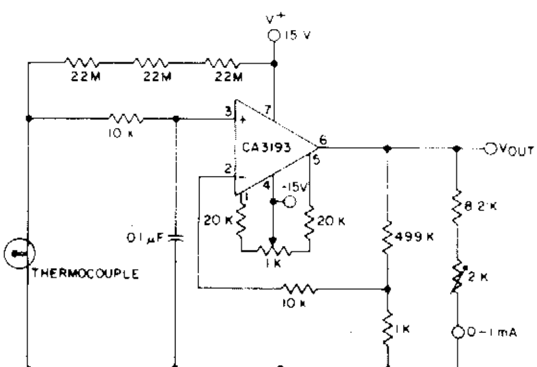
NASA

Fig. 2-14

Circuit Notes

This converter circuit maintains 40 dB of isolation between the input and output halves of a four-wire line while permitting a two-wire line to be connected. A balancing potentiometer, Rg, adjusts the gain of IC2 to null the feed-through from the input to the output. The adjustment is done on the workbench just after assembly by inserting a 1 kHz tone into the four-wire input and setting R_g for minimum output signal. An 82 ohm dummy-load resistor is placed across the two wire terminals.

THERMOCOUPLE AMPLIFIER



ALL RESISTORS ARE 1%

ALL RESISTANCE VALUES ARE IN OHMS

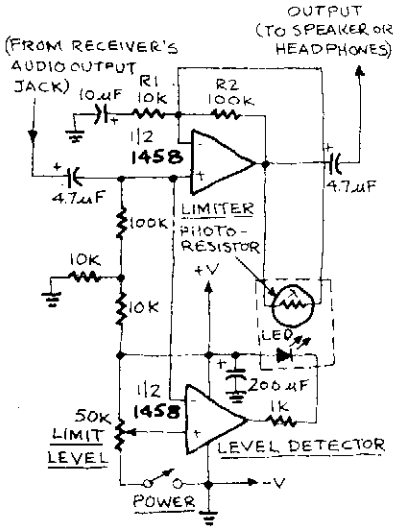
GENERAL ELECTRIC/RCA

Fig. 2-15

Circuit Notes

The circuit uses a CA3193 BiMOS precision op amp to amplify the generated signal 500 times. Three 22-megohm resistors will provide full-scale output if the thermocouple opens.

LOW-DISTORTION AUDIO LIMITER



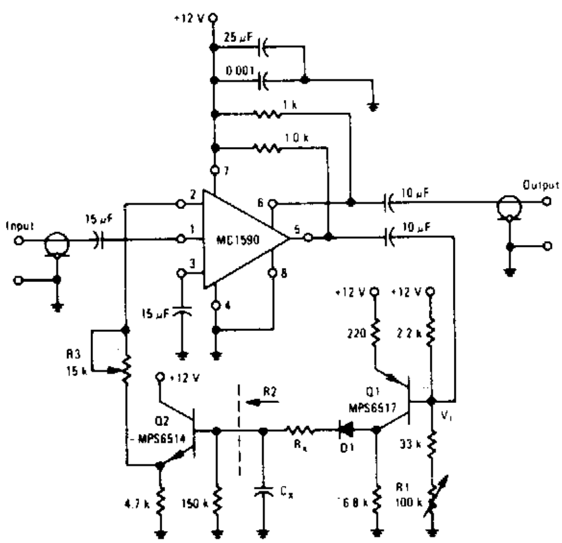
RADIO-ELECTRONICS

Fig. 2-16

Circuit Notes

The level at which the audio limiter comes into action can be set with the LIMIT LEVEL trimmer potentiometer. When that level is exceeded, the output from the LIMITER-DETECTOR half of the op-amp (used as a comparator) turns the LED which causes the resistance of the photoresistor to decrease rapidly. That in turn causes the gain of the LIMITER half of the op-amp to decrease. When the signal drops below the desired limiting level, the LED turns off, the resistance of the photoresistor increases, and the gain of the LIMITER op-amp returns to its normal level—that set by the combination of resistors R1 and R2. A dual-polarity power supply (±12 volts is desirable) is needed for the op-amp.

SPEECH COMPRESSOR



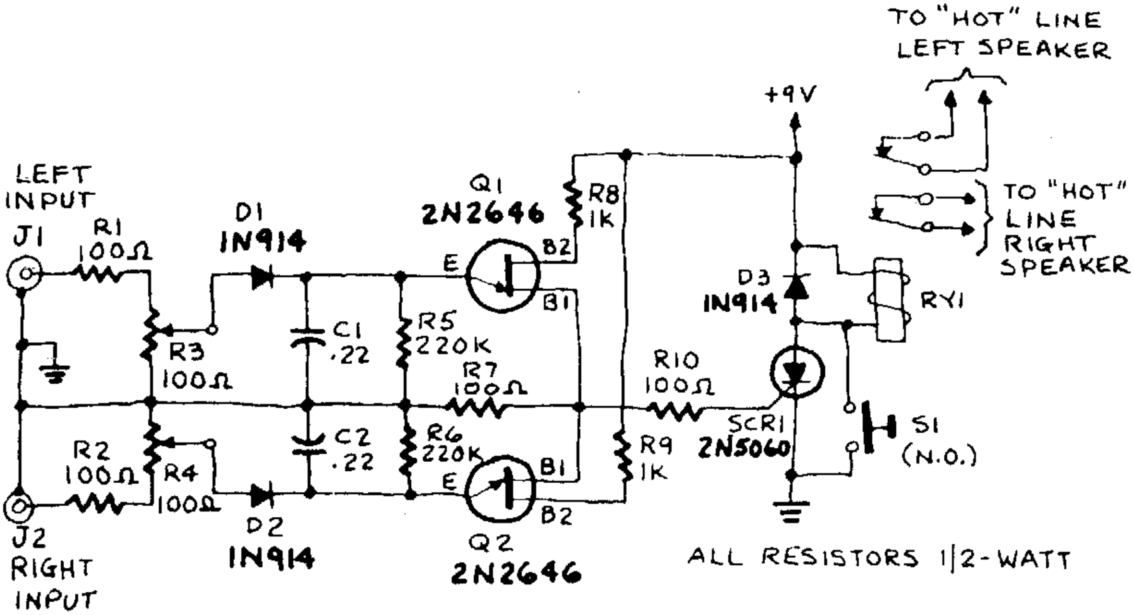
MOTOROLA INC.

Fig. 2-17

Circuit Notes

The amplifier drives the base of a pnp MPS6517 operating common-emitter with a voltage gain of approximately 20. The control R1 varies the quiescent Q point of this transistor so that varying amounts of signal exceed the level V_r. Diode D1 rectifies the positive peaks of Q1's output only when these peaks are greater than $V_r \simeq 7.0$ volts. The resulting output is filtered C_x, R_x. R_x controls the charging time constant or attack time. C_x is involved in both charge and discharge. R2 (150 K, input resistance of the emitter-follower Q2) controls the decay time. Making the decay long and attack short is accomplished by making R_x small and R2 large. (A Darlington emitter-follower may be needed if extremely slow decay times are required.) The emitter-follower Q2 drives the AGC Pin 2 of the MC1590 and reduces the gain. R3 controls the slope of signal compression.

SPEAKER OVERLOAD PROTECTOR



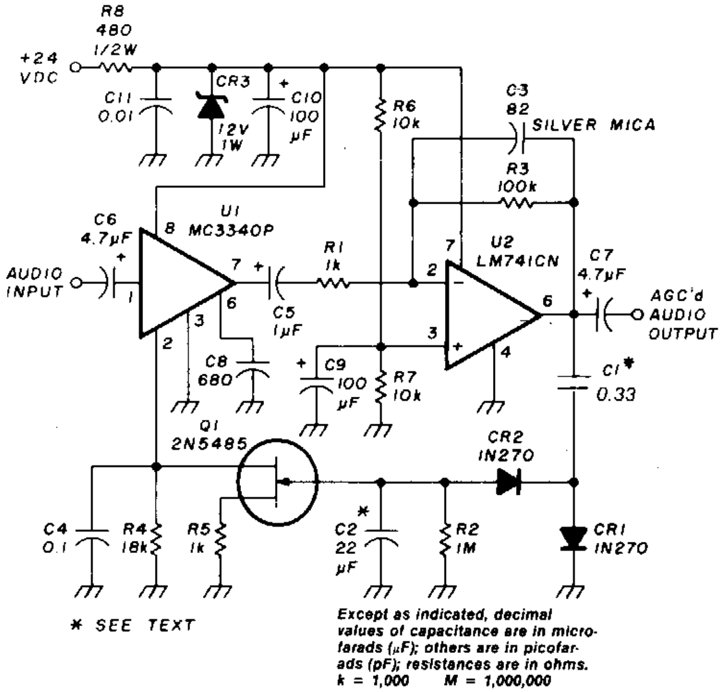
RADIO ELECTRONICS

Fig. 2-18

Circuit Notes

The input to the circuit is taken from the amplifier's speaker-output terminals or jacks. If the right-channel signal is sufficiently large to charge C1 to a potential that is greater than the breakdown voltage of Q1's emitter, a voltage pulse will appear across R7. Similarly, if the left-channel signal is sufficiently large to charge C2 to a potential that is greater than the breakdown voltage of Q2's emitter, a pulse will appear across R7. The pulse across R7 triggers SCR1, a sensitive gate SCR (I_{GT} less than 15 mA where I_{GT} is the gate trigger-current), that latches in a conducting state and energizes RY1. The action of the relay will interrupt both speaker circuits, and the resulting silence should alert you to the problem. Cut back the volume on your amplifier, then press and release S1 to reset the circuit and restore normal operation. The circuit can be adjusted to trip at any level from 15 to 150 watts RMS. To calibrate, deliberately feed an excessive signal to the right input of the speaker protector and adjust R3 until RY1 energizes. Do the same with the left channel, this time adjusting R4. The circuit is now calibrated and ready for use.

AUDIO AUTOMATIC GAIN CONTROL



HAM RADIO

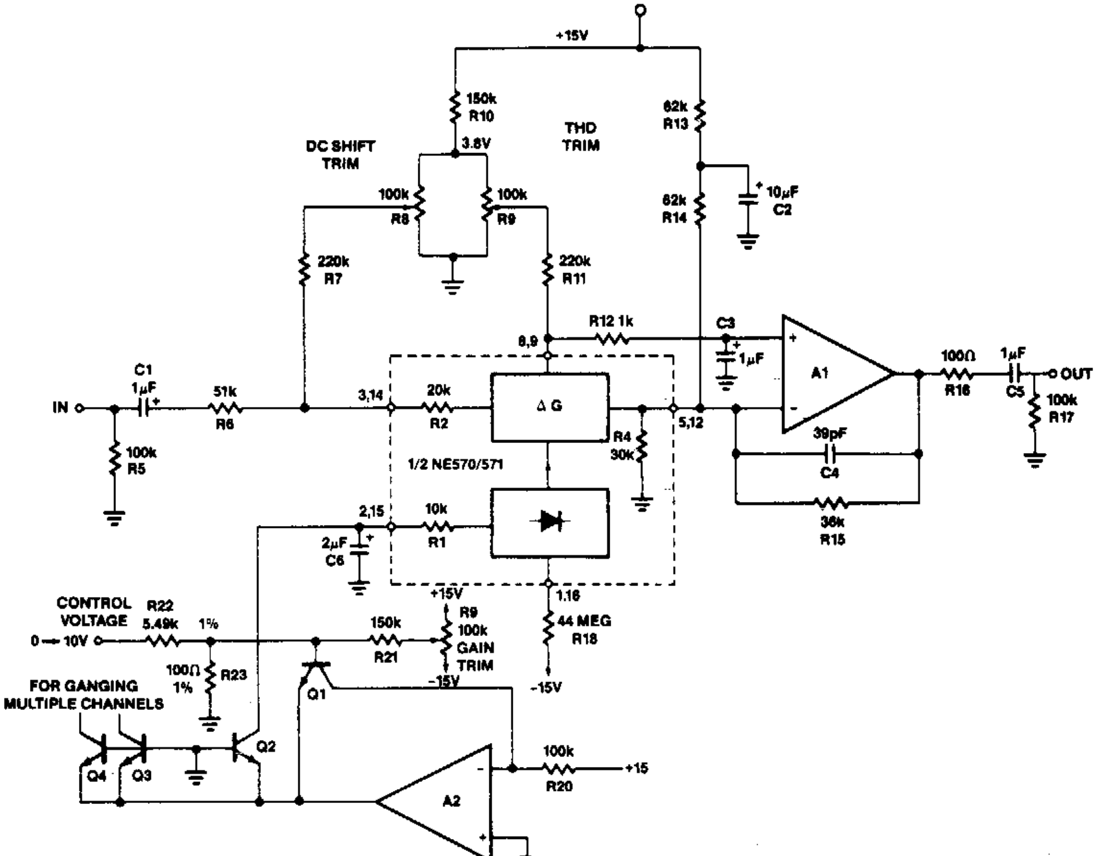
Fig. 2-19

Circuit Notes

An audio signal applied to U1 is passed through to the 741 operational amplifier, U2. After being amplified, the output signal of U2 is sampled and applied to a negative voltage doubler/rectifier circuit composed of diodes CR1 and CR2 along with capacitor C1. The resulting negative voltage is used as a control voltage that is applied to the gate of the 2N5485 JFET Q1. Capacitor C2 and resistor R2 form a smoothing filter for the rectified audio control voltage.

The JFET is connected from pin 2 of the MC3340P to ground through a 1 kilohm resistor. As the voltage applied to the gate of the JFET becomes more negative in magnitude, the channel resistance of the JFET increases causing the JFET to operate as a voltage controlled resistor. The MC3340P audio attenuator is the heart of the AGC. It is capable of 13 dB gain or nearly −80 dB of attenuation depending on the external resistance placed between pin 2 and ground. An increase of resistance decreases the gain achieved through the MC3340P. The circuit gain is not entirely a linear function of the external resistance but approximates such behavior over a good portion of the gain/attenuation range. An input signal applied to the AGC input will cause the gate voltage of the JFET to become proportionally negative. As a result the JFET increases the resistance from pin 2 to ground of the MC3340P causing a reduction in gain. In this way the AGC output is held at a nearly constant level.

VOLTAGE-CONTROLLED ATTENUATOR



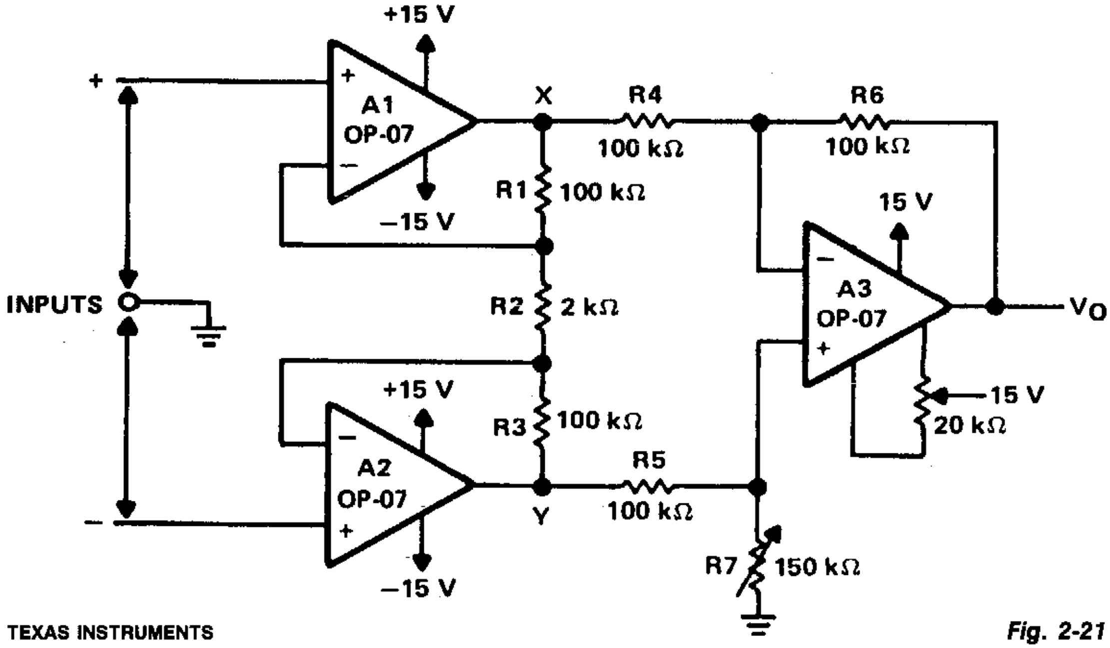
SIGNETICS

Fig. 2-20.

Circuit Notes

Op amp A_2 and transistors Q_1 and Q_2 form the exponential converter generating an exponential gain control current, which is fed into the rectifier. A reference current of 150 μA, (15 V and R_{20} = 100 k), is attenuated a factor of two (6 dB) for every volt increase in the control voltage. Capacitor C_6 slows down gain changes to a 20 ms time constant ($C_6 \times R_1$) so that an abrupt change in the control voltage will produce a smooth sounding gain change. R_{18} ensures that for large control voltages the circuit will go to full attentuation. The rectifier bias current would normally limit the gain reduction to about 70 dB. R_{16} draws excess current out of the rectifier. After approximately 50 dB of attentuation at a −6 dB/V slope, the slope steepens and attenuation becomes much more rapid until the circuit totally shuts off at about 9 V of control voltage. A_1 should be a low-noise high slew rate op amp. R_{13} and R_{14} establish approximately a 0 V bias at A_1's output.

HIGH-INPUT-IMPEDANCE DIFFERENTIAL AMPLIFIER



TEXAS INSTRUMENTS

Fig. 2-21

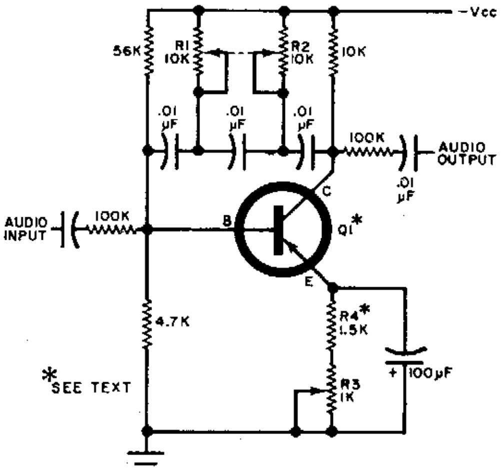
Circuit Notes

Operational amplifiers A1 and A2 are connected in a non-inverting configuration with their outputs driving amplifier A3. Operational amplifier A3 could be called a subtractor circuit which converts the differential signal floating between points X and Y into a single-ended output voltage. Although not mandatory, amplifier A3 is usually operated at unity gain and R4, R5, R6, and R7 are all equal.

The common-mode-rejection of amplifier A3 is a function of how closely the ratio R4:R5 matches the ratio R6:R7. For example, when using resistors with 0.1% tolerance, common-mode rejection is greater than 60 dB. Additional improvement can be attained by using a potentiometer (slightly higher in value than R6) for R7. The potentiometer can be adjusted for the best common-mode rejection. Input amplifiers A1 and A2 will have some differential gain but the common-mode input voltages will experience only unity gain. These voltages will not appear as differential signals at the input of amplifier A3 because, when they appear at equal levels on both ends of resistor R2, they are effectively canceled.

This type of low-level differential amplifier finds widespread use in signal processing. It is also useful for dc and low-frequency signals commonly received from a transducer or thermocouple output, which are amplified and transmitted in a single-ended mode. The amplifier is powered by ±15 V supplies. It is only necessary to null the input offset voltage of the output amplifier A3.

AUDIO Q-MULTIPLIER



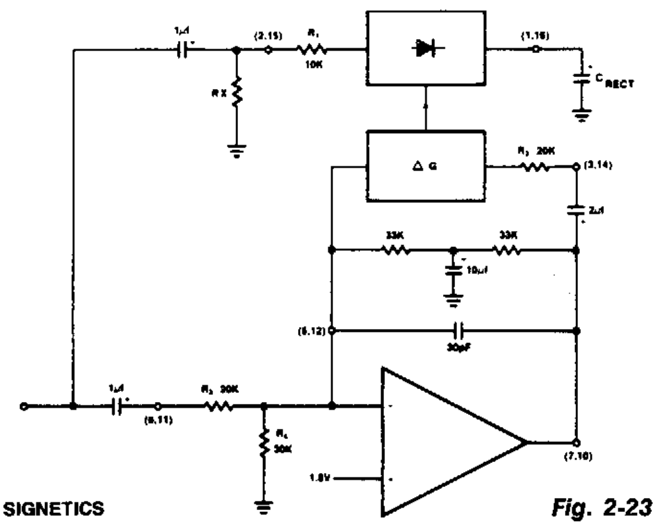
POPULAR ELECTRONICS

Fig. 2-22

Circuit Notes

This circuit is for selective tuning between two closely spaced audio tones. The selective frequency is dependent on the value of capacitors and resistors in the feedback circuit between the collector and base of Q1. With the values shown, the frequency can be "tuned" a hundred cycles or so around 650 Hz. R1 and R2 must be ganged. Emitter potentiometer R3 determines the sharpness of response curve. Any transistor having a beta greater than 50 can be used. Select a value for R4 so that the circuit will not oscillate when R3 is set for minimum bandwidth (sharpest tuning).

AUTOMATIC LEVEL CONTROL



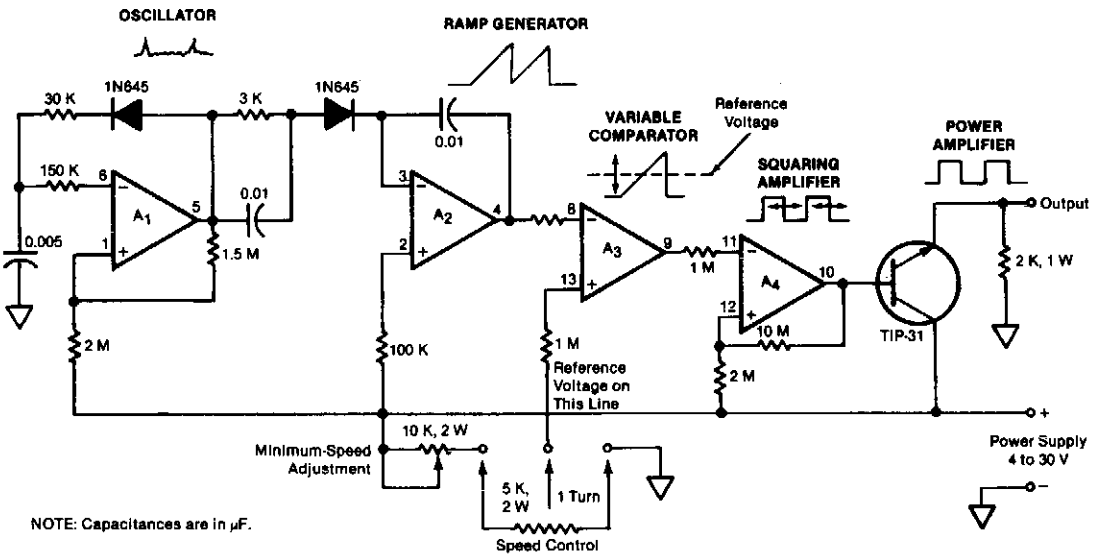
SIGNETICS

Fig. 2-23

Circuit Notes

The NE570 can be used to make a very high performance ALC compressor, except that the rectifier input is tied to the input. This makes gain inversely proportional to input level so that a 20 dB drop in input level will produce a 20 dB increase in gain. The output will remain fixed at a constant level. As shown, the circuit will maintain an output level of ±1 dB for an input range of +14 to −43 dB at 1 kHz. Additional external components will allow the output level to be adjusted.

PULSE-WIDTH PROPORTIONAL-CONTROLLER CIRCUIT



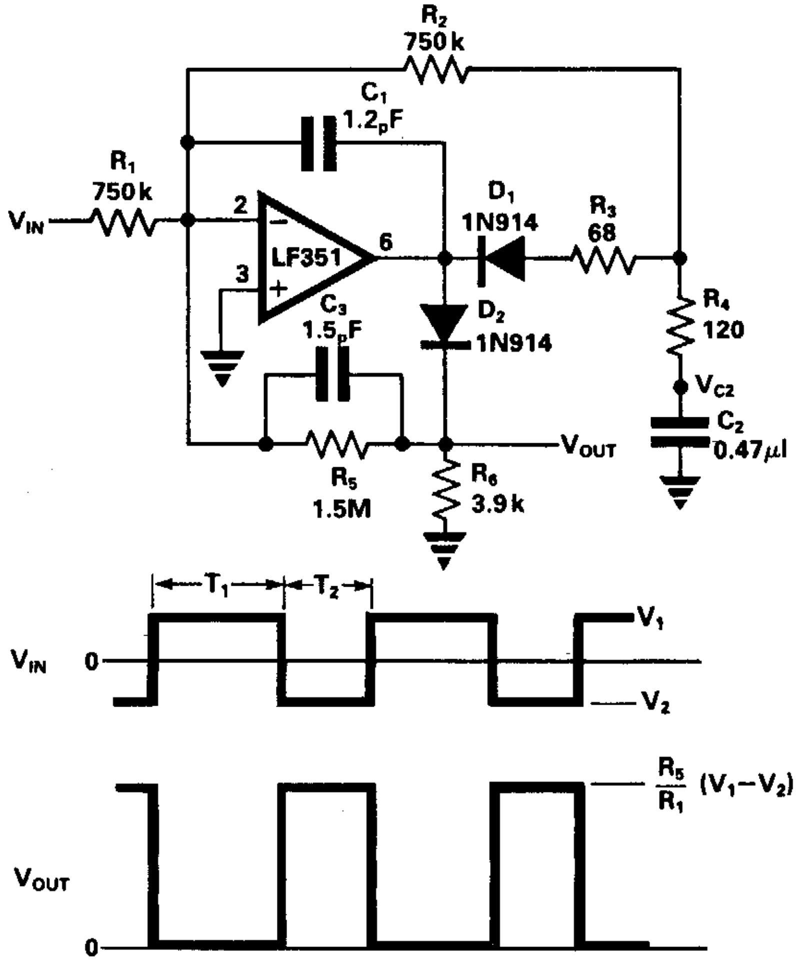
NASA

Fig. 2-24

Circuit Notes

The quad operational amplifier circuit yields full 0 to 100 percent pulse-width control. The controller uses an LM3900 that requires only a single supply voltage of 4 to 30 V. The pulse-repetition rate is set by a 1 kHz oscillator that incorporates amplifier A_1. The oscillator feeds ramp generator A_2, which generates a linear ramp voltage for each oscillator pulse. The ramp signal feeds the inverting input of comparator A_3; the speed-control voltage feeds the noninverting input. Thus, the output of the comparator is a 1 kHz pulse train, the pulse width of which changes linearly with the control voltage. The control voltage can be provided by an adjustable potentiometer or by an external source of feedback information such as a motor-speed sensing circuit. Depending on the control-voltage setting, the pulse duration can be set at any value from zero (for zero average dc voltage applied to the motor) to the full pulse-repetition period (for applied motor voltage equal to dc power-supply voltage). An amplifier stage (A_4) with a gain of 10 acts as a pulse-squaring circuit. A TIP-31 medium-power transistor is driven by A_4 and serves as a separate power-amplifier stage.

OP AMP CLAMPING



ELECTRONIC ENGINEERING *Fig. 2-25*

Circuit Notes

The circuit clamps the most positive value of the input pulse signal to the zero base level. Additionally, the circuit inverts and amplifies the input signal by the factor of R_5/R_1. The waveforms are shown in the bottom of Fig. 2-24.

3

Analog-to-Digital Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SUCCESSIVE APPROXIMATION A/D CONVERTERS

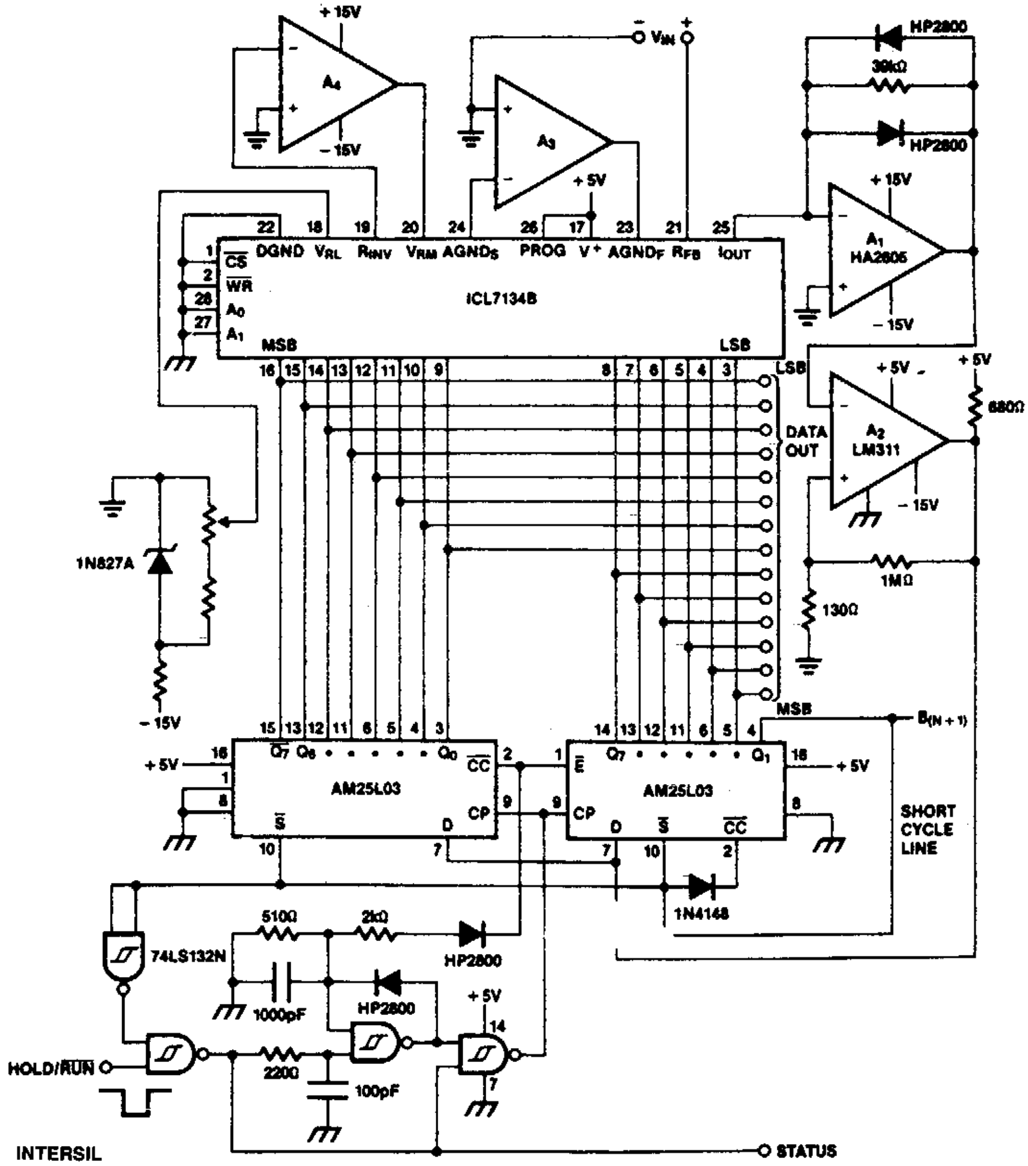


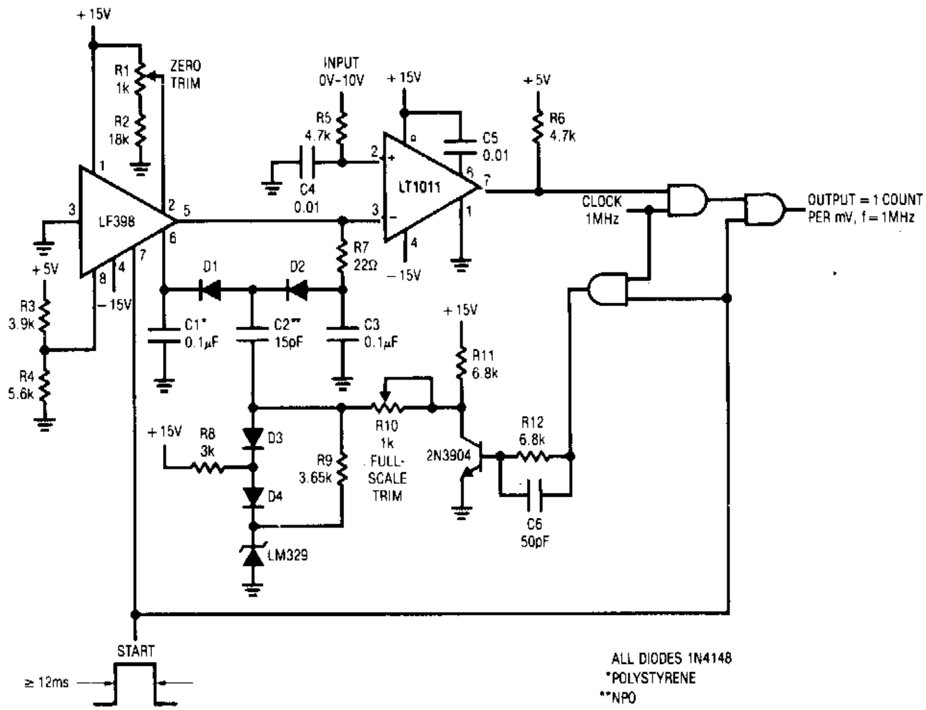
Fig. 3-1

SUCCESSIVE APPROXIMATION A/D CONVERTERS , Continued.

Circuit Notes

The ICL7134B-based circuit is for a bipolar-input high-speed A/D converter, using two AM25L03s to form a 14-bit successive approximation register. The comparator is a two-stage circuit with an HA2605 front-end amplifier, used to reduce settling time problems at the summing node (see A020). Careful offset-nulling of this amplifier is needed, and if wide temperature range operation is desired, an auto-null circuit using an ICL7650 is probably advisable (see A053). The clock, using two Schmitt trigger TTL gates, runs at a slower rate for the first 8 bits, where settling-time is most critical than for the last 6 bits. The short-cycle line is shown tied to the 15th bit; if fewer bits are required, it can be moved up accordingly. The circuit will free-run if the HOLD/RUN input is held low, but will stop after completing a conversion if the pin is high at that time. A low-going pulse will restart it. The STATUS output indicates when the device is operating, and the falling edge indicates the availability of new data. A unipolar version can be constructed by typing the MSB (D13) on an ICL7134U to pin 14 on the first AM25L03, deleting the reference inversion amplifier A4, and tying V_{RFM} to V_{RFL}.

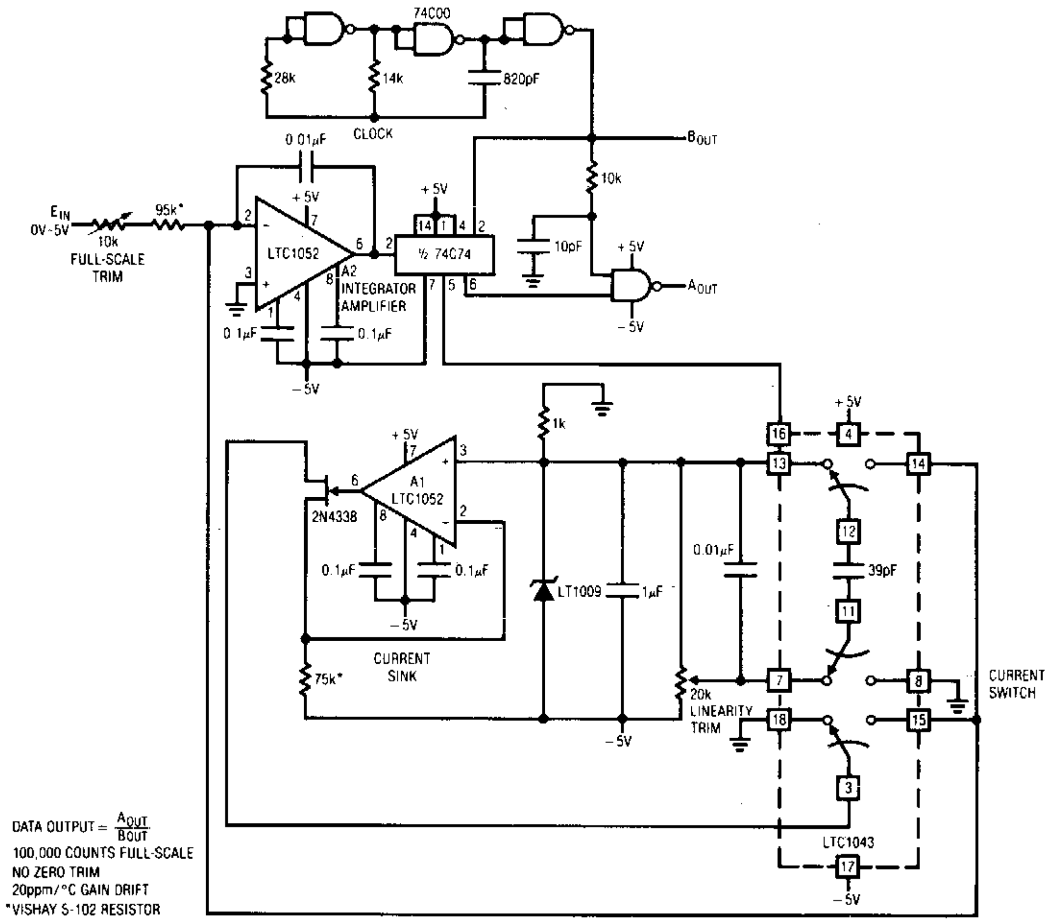
4 DIGIT (10,000 COUNT) A/D CONVERTER



LINEAR TECHNOLOGY CORP.

Fig. 3-2

16-BIT A/D CONVERTER



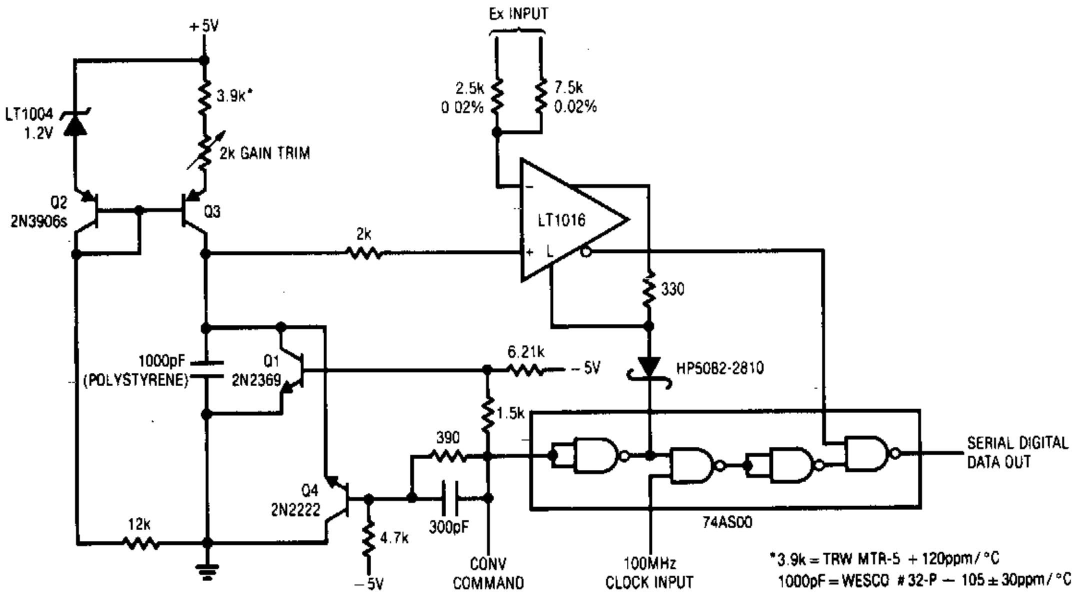
LINEAR TECHNOLOGY

Fig. 3-3

Circuit Notes

The A/D converter, made up of A2, a flip-flop, some gates and a current sink, is based on a current balancing technique. Once again, the chopper-stabilized LTC1052's 50 nV/°C input drift is required to eliminate offset errors in the A/D.

INEXPENSIVE, FAST 10-BIT SERIAL OUTPUT A/D



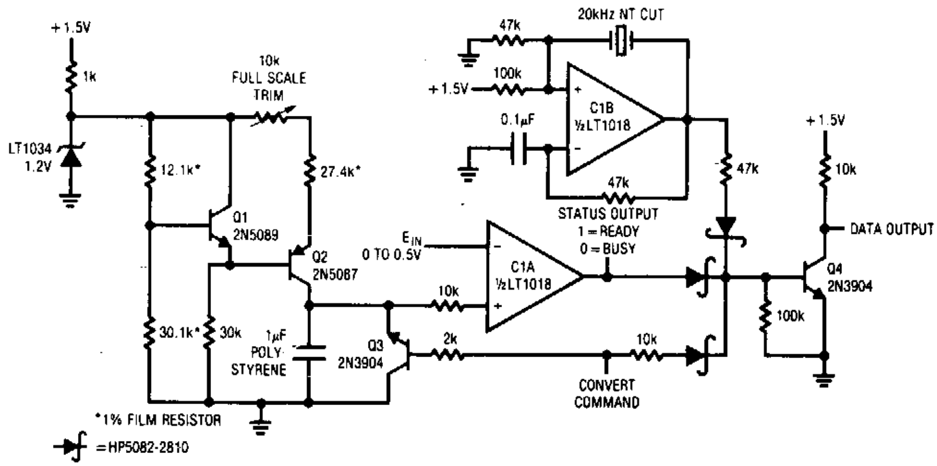
LINEAR TECHNOLOGY CORPORATION

Fig. 3-4

Circuit Notes

Everytime a pulse is applied to the convert command input, Q1 resets the 1000 pF capacitor to 0 V. This resetting action takes 200 ns of the falling edge of the convert command pulse, the capacitor begins to charge linearly. In precisely 10 microseconds, it charges to 2.5 V. The 10 microseconds ramp is applied to the LT1016's positive input. The LT1016 compares the ramp to Ex, the unknown, at its negative input. For a 0 V - 2.5 V range, Ex is applied to the 2.5 k ohm resistor. From a 0 V - 10 V range, the 2.5 k ohm resistor is grounded and Ex is applied to the 7.5 k ohm resistor. Output of the LT1016 is a pulse whose width is directly dependent on the value of Ex. This pulse width is used to gate a 100 MHz clock. The 100 MHz clock pulse bursts that appear at the output are proportional to Ex. For a 0 V - 10 V input, 1024 pulses appear at full-scale, 512 at 5.00 V, etc.

10-BIT A/D CONVERTER



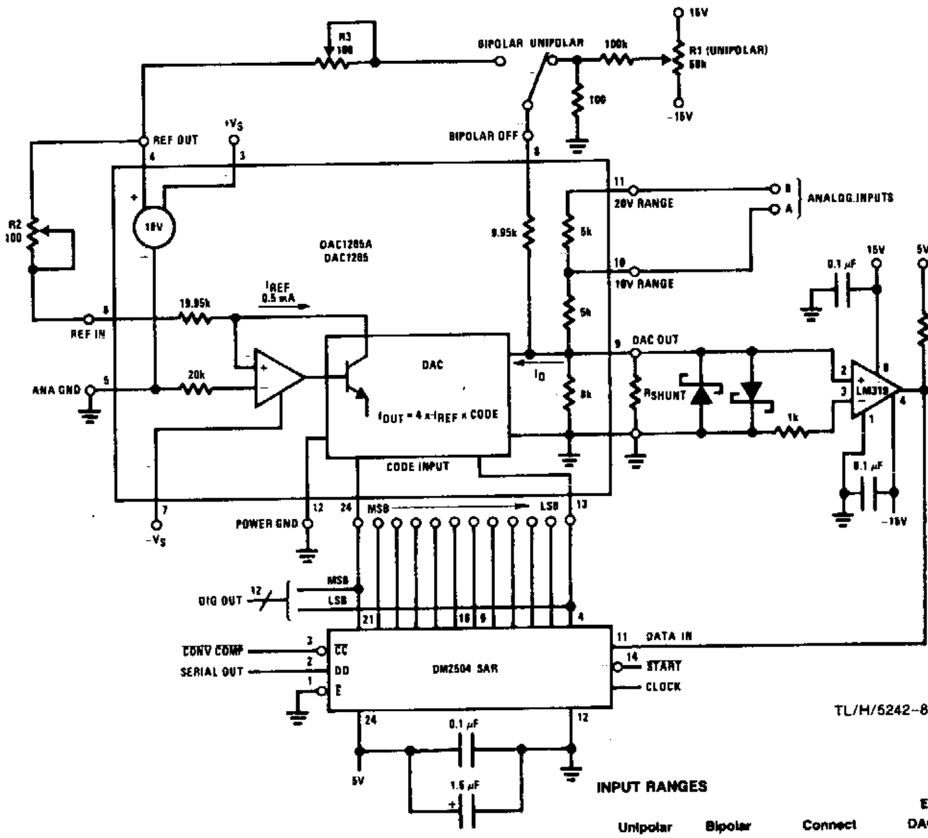
LINEAR TECHNOLOGY CORPORATION

Fig. 3-5

Circuit Notes

The converter has a 60 ms conversion time, consumes 460 μA from its 1.5 V supply and maintains 10 bit accuracy over a 15°C to 35°C temperature range. A pulse applied to the convert command line causes Q3, operating in inverted mode, to discharge through the 10 kΩ diode path, forcing its collector low. Q3's inverted mode switching results in a capacitor discharge within 1 mV of ground. During the time the ramps' value is below the input voltage, CIA's output is low. This allows pulses from C1B, a quartz stabilized oscillator, to modulate Q4. Output data appears at Q4's collector. When the ramp crosses the input voltages value C1A's output goes high, biasing Q4 and output data ceases. The number of pulses at the output is directly proportional to the input voltage. To calibrate apply 0.5 V to the input and trim the 10 kΩ potentiometer for exactly 1000 pulses out each time the convert command line is pulsed.

HIGH SPEED 12-BIT A/D CONVERTER



INPUT RANGES

Unipolar	Bipolar	Connect	Equiv. DAC Z_{OUT}
0 to 5	±2.5	Input to A B to DAC OUT	1.60 kΩ
0 to 10	±5	Input to A	2.35 kΩ
0 to 20	±10	Input to B	3.08 kΩ

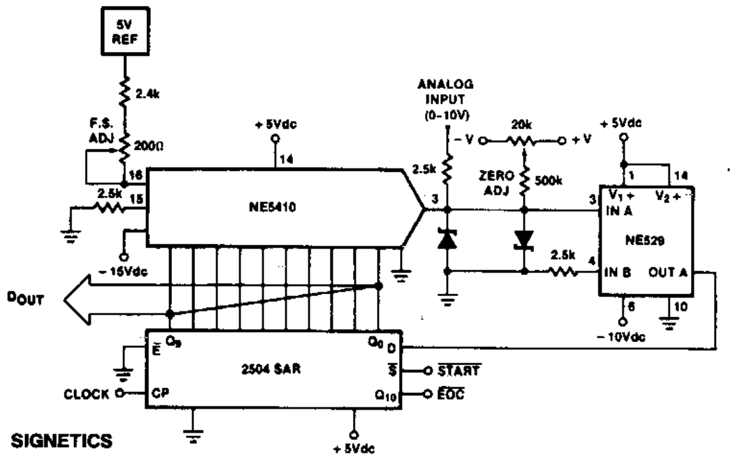
NATIONAL SEMICONDUCTOR CORP.

Fig. 3-6

Circuit Notes

This system completes a full 12-bit conversion in 10 µs unipolar or bipolar. This converter will be accurate to ± ½ LSB of 12 bits and have a typical gain TC of 10 ppm/°C. In the unipolar mode, the system range is 0 V to 9.9976 V, with each bit having a value of 2.44 mV. For the true conversion accuracy, an A/D converter should be trimmed so that given bit code output results from input levels from ½ LSB below to ½ LSB above the exact voltage which that code represents. Therefore, the converter zero point should be trimmed with an input voltage of 1.22 mV; trim R1 until the LSB just begins to appear in the output code (all other bits "0"). For full-scale, use an input voltage of 9.9963 V (10 V-1 LSB-½ LSB); then trim R2 until the LSB just begins to appear (all other bits "1"). The bipolar signal range is −5.0 V to 4.9976 V. Bipolar offset trimming is done by applying a −4.9988 V input signal and trimming R3 for the LSB transition (all other bits "0"). Full-scale is set by applying 4.9963 V and trimming R2 for the LSB transition (all other bits "1").

SUCCESSIVE APPROXIMATION A/D CONVERTER



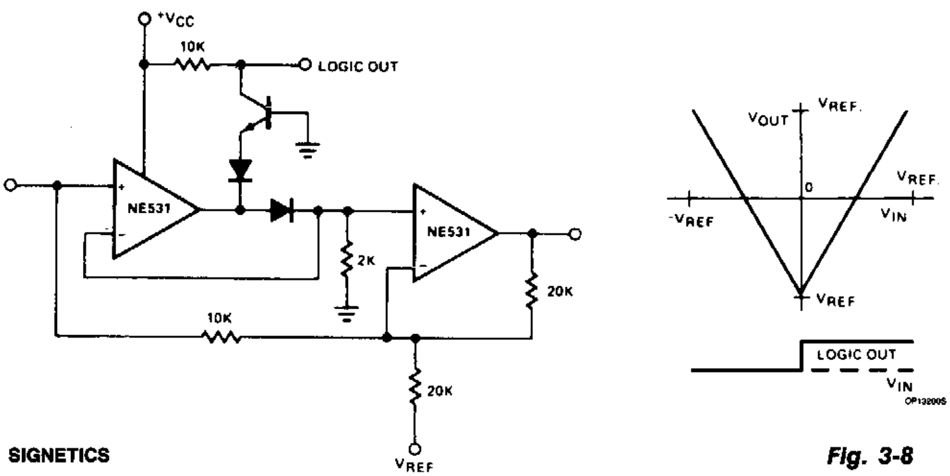
SIGNETICS

Fig. 3-7

Circuit Notes

The 10-bit conversion time is 3.3 μs with a 3 MHz clock. This converter uses a 2504 12-bit successive approximation register in the short cycle operating mode where the end of conversion signal is taken from the first unused bit of the SAR (Q_{10}).

CYCLIC A/D CONVERTER



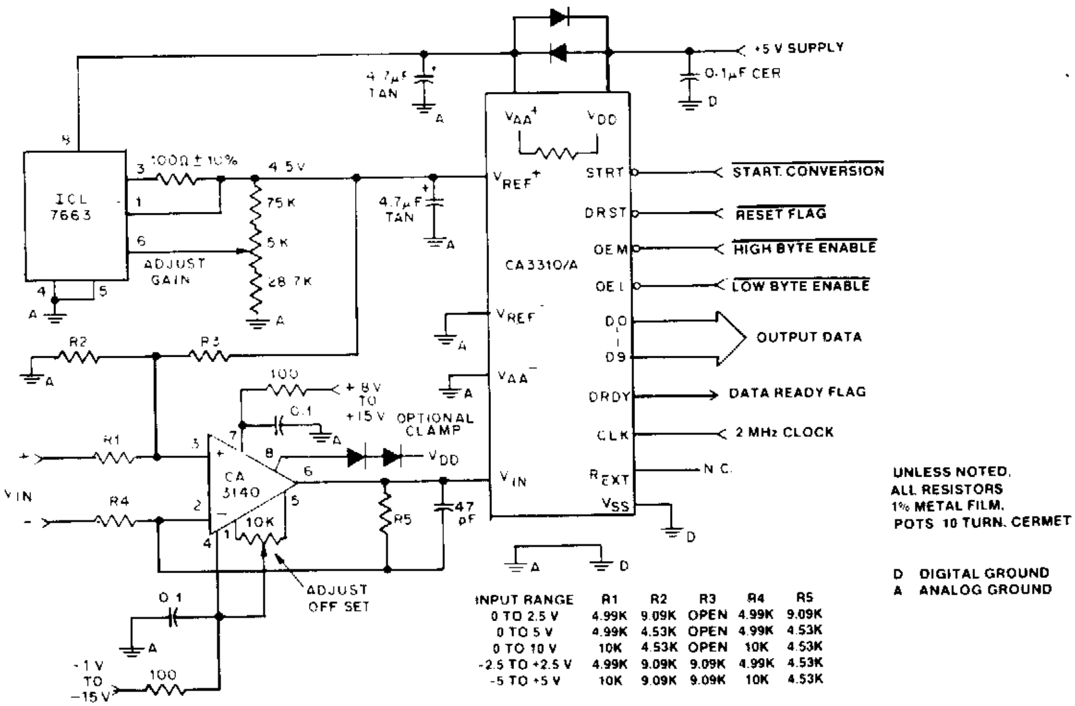
SIGNETICS

Fig. 3-8

Circuit Notes

The cyclic converter consists of a chain of identical stages, each of which senses the polarity of the input. The stage then subtracts V_{REF} from the input and doubles the remainder if the polarity was correct. The signal is full-wave rectified and the remainder of $V_{IN} - V_{REF}$ is doubled. A chain of these stages gives the gray code equivalent of the input voltage in digitized form related to the magnitude of V_{REF}. Possessing high potential accuracy, the circuit using NE531 devices settles in 5 μs.

DIFFERENTIAL INPUT A/D SYSTEM



INPUT RANGE	R1	R2	R3	R4	R5
0 TO 2.5 V	4.99K	9.09K	OPEN	4.99K	9.09K
0 TO 5 V	4.99K	4.53K	OPEN	4.99K	4.53K
0 TO 10 V	10K	4.53K	OPEN	10K	4.53K
-2.5 TO +2.5 V	4.99K	9.09K	9.09K	4.99K	4.53K
-5 TO +5 V	10K	9.09K	9.09K	10K	4.53K

GENERAL ELECTRIC/RCA

Fig. 3-9

Circuit Notes

Using a CA3140 BiMOS op amp provides good slewing capability for high bandwidth input signals, and can quickly settle energy that the CA3310 outputs at its V_{IN} terminal. The CA3140 can also drive close to the negative supply rail. If system supply sequencing or an unknown input voltage is likely to cause the op amp to drive above the V_{DD} supply, a diode clamp can be added from pin 8 of the op amp to the V_{DD} supply.

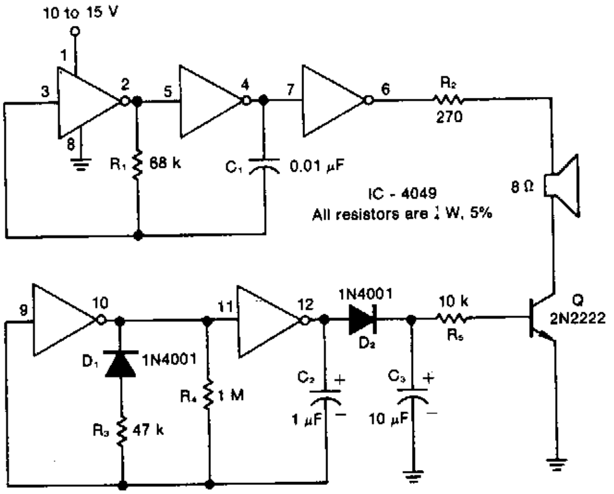
4

Annunciators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low-Cost Chime Circuit
Electronic Bell
Sliding-Tone Doorbell

LOW-COST CHIME CIRCUIT



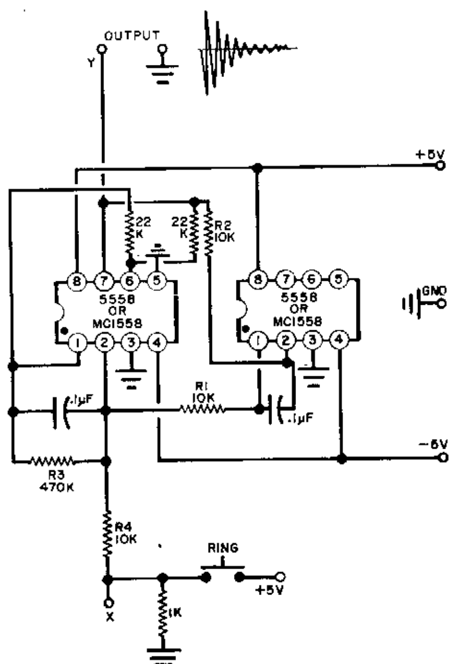
Circuit Notes

Resistor R1, capacitor C1, and two inverters form a square wave generator, which produces the basic tone. The generator is followed by an inverter that acts as both a buffer and a driver for the speaker. Resistor R2, which has a minimum value of 100 ohms, limits the current and controls the volume. Diode D1, capacitor C2, resistors R3 and R4, and two inverters create the pulse generator that determines the turn-on and decay times of the chime. The decay circuit—formed by D2, C3, R5, and Q—reduces the amplitude of the chime tone exponentially as a function of time.

ELECTRONIC DESIGN

Fig. 4-1

ELECTRONIC BELL



Circuit Notes

Coarse frequency is controlled by the capacitors which must be kept nearly identical in value to each other for best results. Fine tuning is accomplished with R1 and R2. The decay time is controlled by R3.

POPULAR ELECTRONICS

Fig. 4-2

SLIDING-TONE DOORBELL

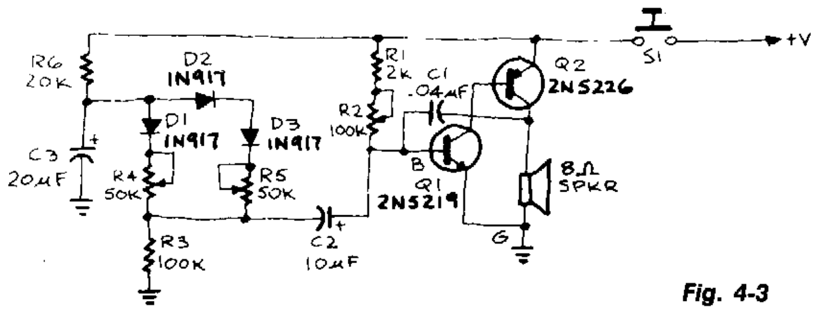


Fig. 4-3

RADIO ELECTRONICS

Circuit Notes

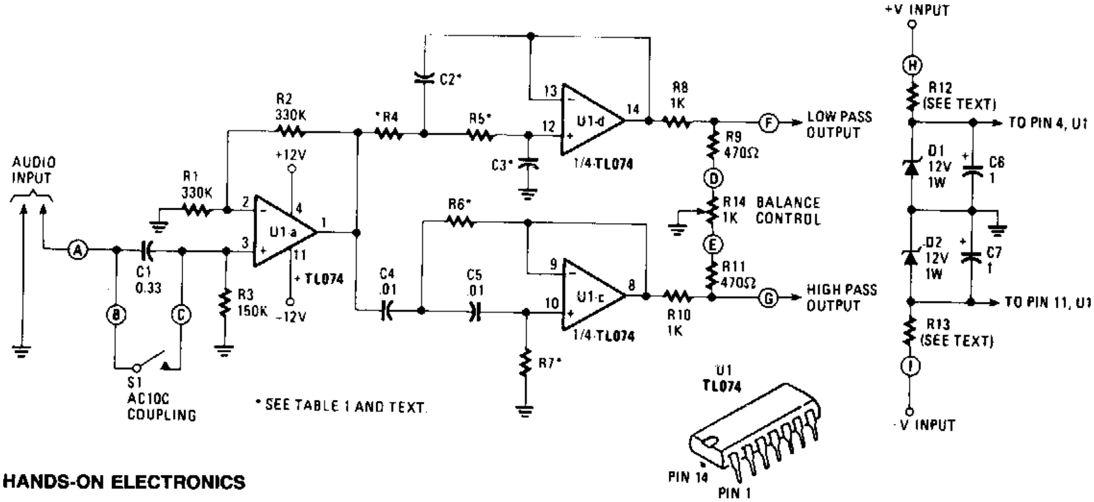
When the doorbell is pushed, you'll hear a low tone that will "slide up" to a higher frequency. The frequency of the AF oscillator is determined by coupling capacitor, C1 and the value of the resistance connected between the base of Q1 and ground. That resistance, R_{BG} is equal to (R1 + R2) R3. First, assume that S1 is closed and R2 has been adjusted to produce a pleasant, low-frequency tone. Capacitor C3 will charge through R6 until it reaches such a voltage that it will cause diode D1 to conduct. When that happens, the value of R_{BG} is paralleled by R4. Thus, because the total resistance R_{BG} decrease, the output tone slides up in frequency. Capacitor C3 will continue to charge until the voltage across D2 and D3 causes those diodes to conduct. Then R_{BG} is paralleled also by R5, the total resistance again decreases, and the oscillator's frequency again increases.

5

Audio Mixers, Crossovers and Distribution Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ELECTRONIC CROSSOVER CIRCUIT



HANDS-ON ELECTRONICS

Fig. 5-1

Circuit Notes

An audio source, such as a mixer, preamplifier, equalizer, or recorder, is fed to the Electronic Crossover Circuit's input. That signal is either ac- or dc-coupled, depending on the setting of switch S1, to the non-inverting input of buffer-amplifier U1a, one section of a quad, BIFET, low-noise TL074 op amp made by Texas Instruments. That stage has a gain of 2, and its output is distributed to both a lowpass filter made by R4, R5, C2, C3, and op-amp U1d, and a highpass filter made by R6, R7, C4, C5, and op amp U1c. Those are 12 dB/octave Butterworth-type filters. The Butterworth filter response was chosen because it gives the best compromise between damping and phase shift. Values of capacitors and resistors will vary with the selected crossover at which your unit will operate. The filter's outputs are fed to a balancing network made by R8, R9, R10, R11 and balance potentiometer R14. When the potentiometer is at its mid-position, there is unity gain for the passbands of both the high and low filters. Dc power for the Electronic Crossover Circuit is regulated by R12, R13, D1, and D2, and decoupled by C6 and C7.

SOUND MIXER/AMPLIFIER

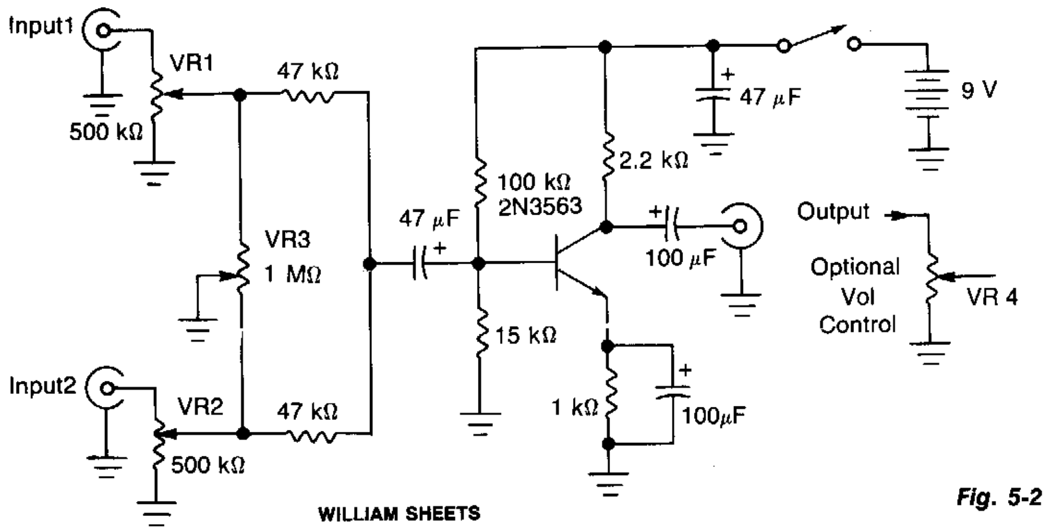


Fig. 5-2

Circuit Notes

Both input signals can be independently controlled by VR1 and VR2. The balance control VR3 is used to fade out one signal while simultaneously fading in the other. The transistor provides gain and the combined output signal level is controlled by VR4 (optional).

MICROPHONE MIXER

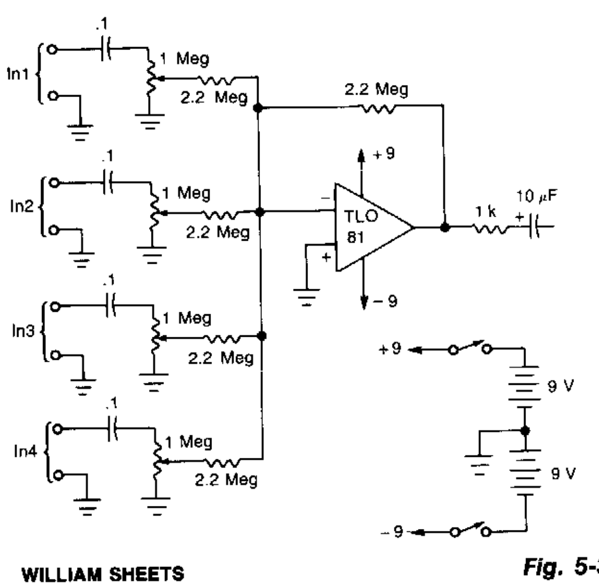


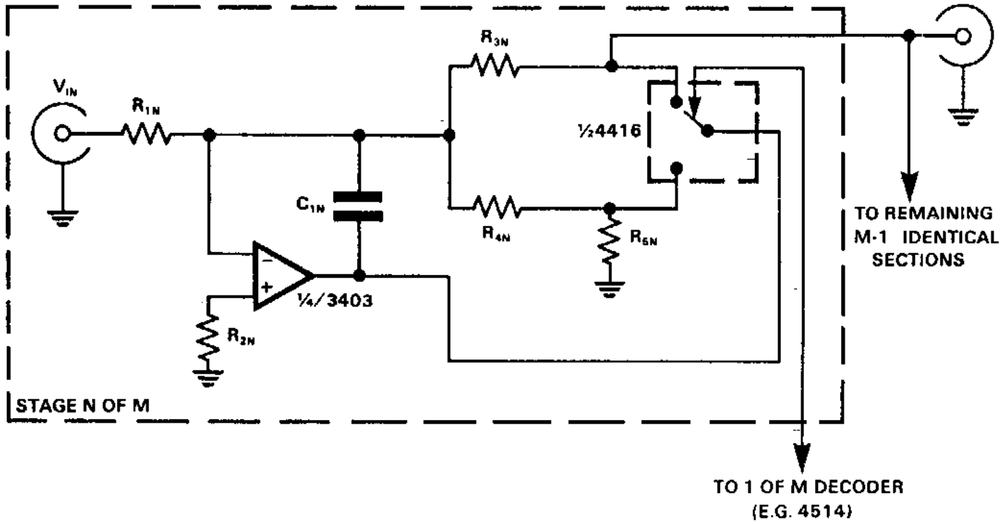
Fig. 5-3

Circuit Notes

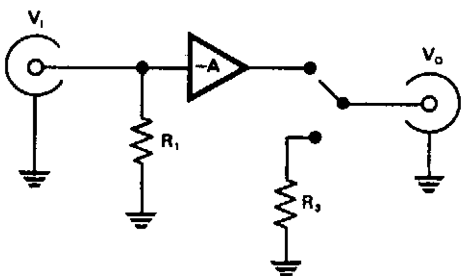
A TL081 op amp is used as a high-to-low impedance converter and signal mixer. The input impedance is approximately 1 megohm and the output impedance is about 1 kilohm. Two 9-volt batteries are used as the power source. Battery life should be several hundred hours with alkaline batteries.

LOW DISTORTION INPUT SELECTOR FOR AUDIO USE

LOW DISTORTION INPUT SELECTOR



EQUIVALENT CIRCUIT OF EACH STAGE:



$R_{3N} = R_{4N} = AR_{1N}$

$R_{2N} = (R_{1N} + R_s) // R_{3N} // (R_{4N} + R_{5N})$

$\frac{1}{2\pi f_{MAX}} \gg R_{3N} C_{1N} \gg \tau_s$

$R_{5N} = \frac{1}{R_L^{-1} + \sum_{i=1}^{M} (R_{3i})^{-1}}$

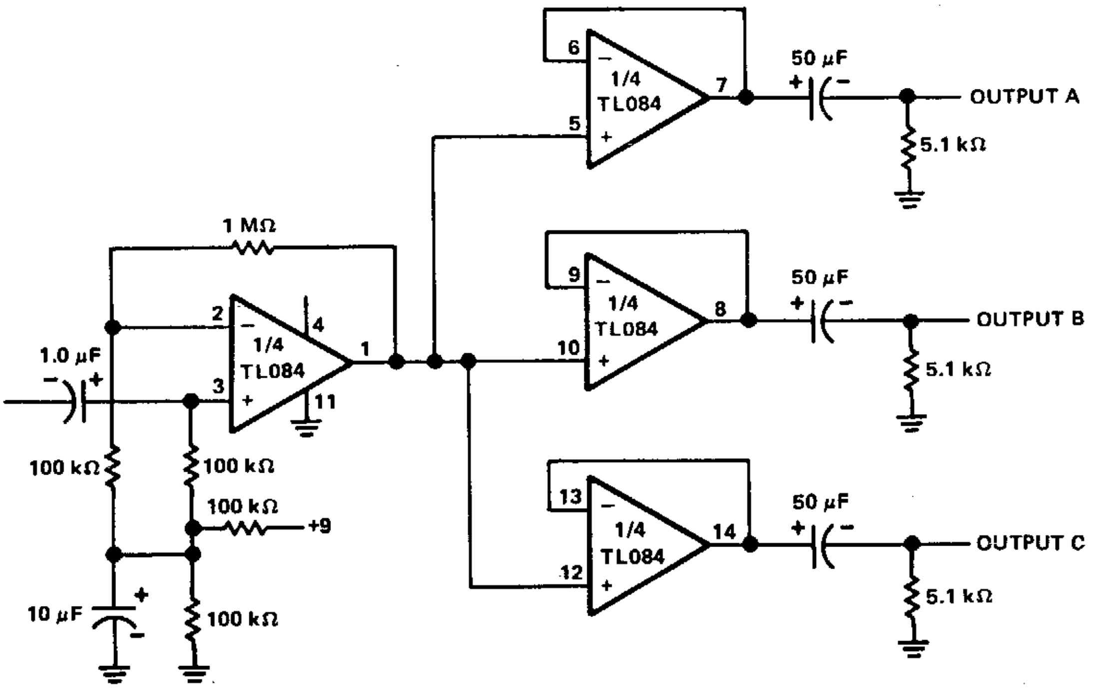
ELECTRONIC ENGINEERING

Fig. 5-4

Circuit Notes

CMOS switches are used directly to select inputs in audio circuits, this can introduce unacceptable levels of distortion, but if the switch is included in the feedback network of an op amp, the distortion due to the switch can be almost eliminated. The circuit uses a 4416 CMOS switch, arranged as two independent SPDT switches. If switching transients are unimportant, R5 and C1 can be omitted, and R4 can be shorted out. However, a feedback path must be maintained, even when a channel is switched out, in order to keep the inverting input of the op amp at ground potential, and prevent excessive crosstalk between channels.

AUDIO DISTRIBUTION AMPLIFIER



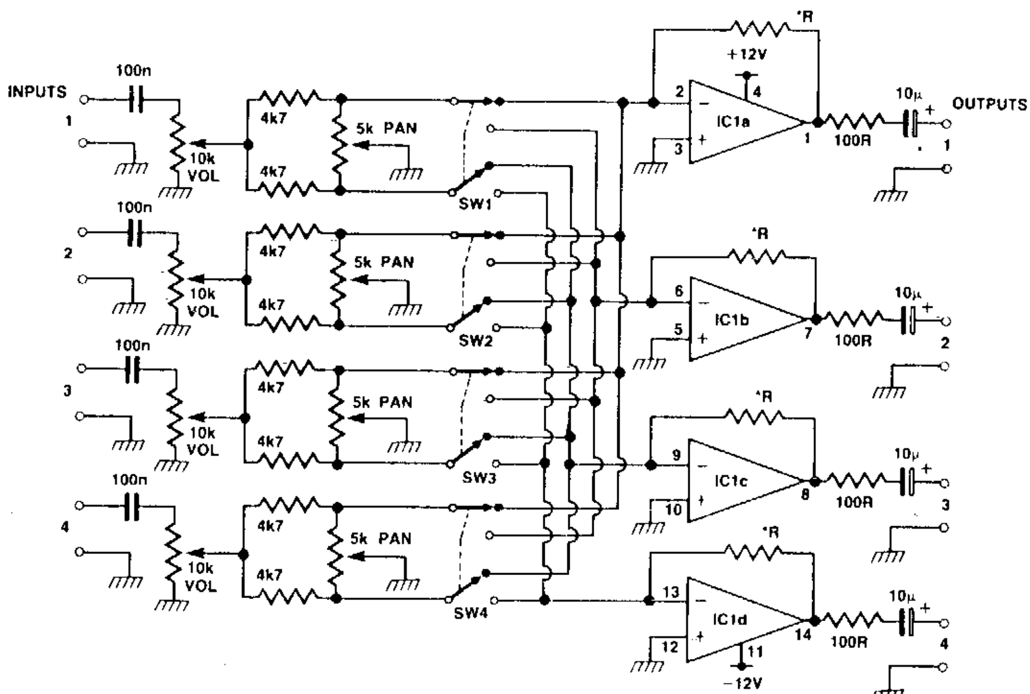
TEXAS INSTRUMENTS

Fig. 5-5

Circuit Notes

The three channel output distribution amplifier uses a single TL084. The first stage is capacitively coupled with a 1.0 μF electrolytic capacitor. The inputs are at ½ V_{CC} rail or 4.5 V. This makes it possible to use a single 9 V supply. A voltage gain of 10 (1 M ohm/100 k ohm) is obtained in the first stage, and the other three stages are connected as unity-gain voltage followers. Each output stage independently drives an amplifier through the 50 μF output capacitor to the 5.1 k ohm load resistor. The response is flat from 10 Hz to 30 kHz.

FOUR CHANNEL FOUR TRACK MIXER



*Note: Choose R to give appropriate gain.

ELECTRONICS TODAY INTERNATIONAL

Fig. 5-6

Circuit Notes

This circuit can be used as a stereo mixer as well as a four track. The quad op-amp IC gives a bit of gain for each track. The pan control allows panning between tracks one and two with the switch in the up position, and with the switch in the down position, it makes possible panning between tracks three and four. Extra channels can be added. A suitable op amp for IC1 is TL074 or similar.

6

Audio Signal Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AUTO FADE

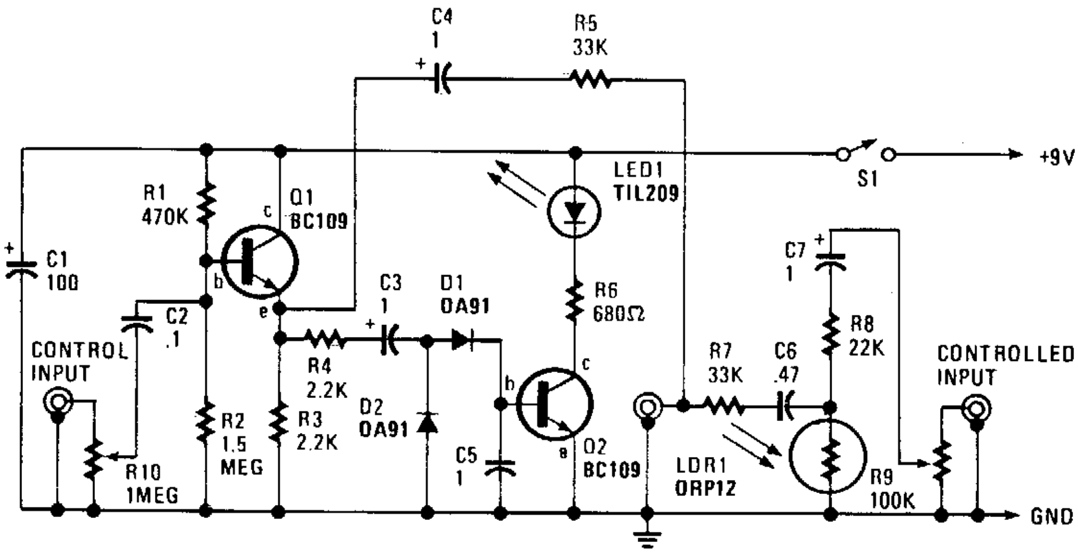


Fig. 6-1

Circuit Notes

The automatic fader drops the level of the background music when the narration comes up. The control input goes through R10, a preset audio level control, to the input of an emitter-follower buffer stage (Q1). The buffer offers a high input impedance and makes sure that the source impedance is low enough to drive the rectifier and smoothing circuit, which consist of D1, D2, and C5. The smoothed output drives a simple LED circuit. R8 and LDR1 form an input attenuator across which the output is fed via C6 and C7 to the output jack. The output at the emitter of Q1 couples to this socket through C4 and R5. R5 and R7 are a passive mixer. With 200 mV or less at the input, there isn't sufficient voltage across C5 to make Q2 turn on. Over 200 mV, Q2 does turn on to a limit, and the LED gets power. That makes the LDR's resistance fall, and signal loss through the attenuator increases. Increase the input to 350 mV rms, and you get a signal reduction of better than 20 dB.

TRANSISTOR HEADPHONE AMPLIFIER

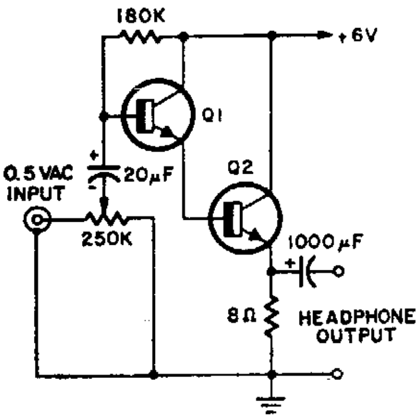
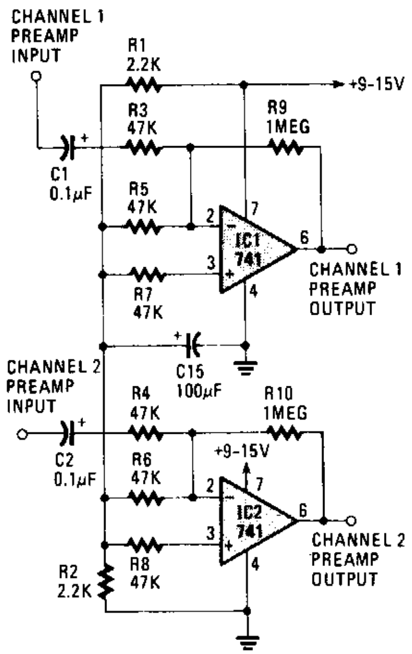


Fig. 6-2

RADIO ELECTRONICS

STEREO PREAMPLIFIER



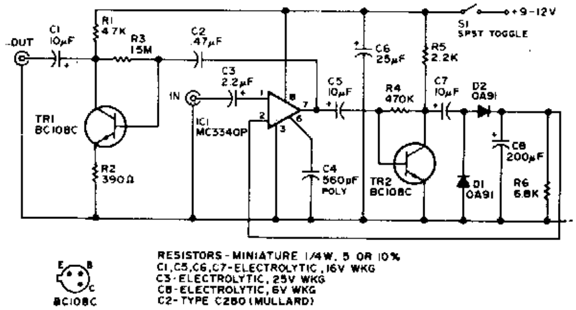
Circuit Notes

The circuit provides better than 20-dB gain in each channel. A better op-amp type will give a better noise figure and bandpass. In this circuit the roll-off is acute at 20,000 Hertz.

HANDS-ON ELECTRONICS

Fig. 6-2

AUDIO COMPRESSOR



73 MAGAZINE

Circuit Notes

A MC3340P is used as a variable gain amplifier. The output of TR2 is rectified and controls the gain of IC1.

Fig. 6-3

MICROPOWER HIGH-INPUT-IMPEDANCE 20-dB AMPLIFIER

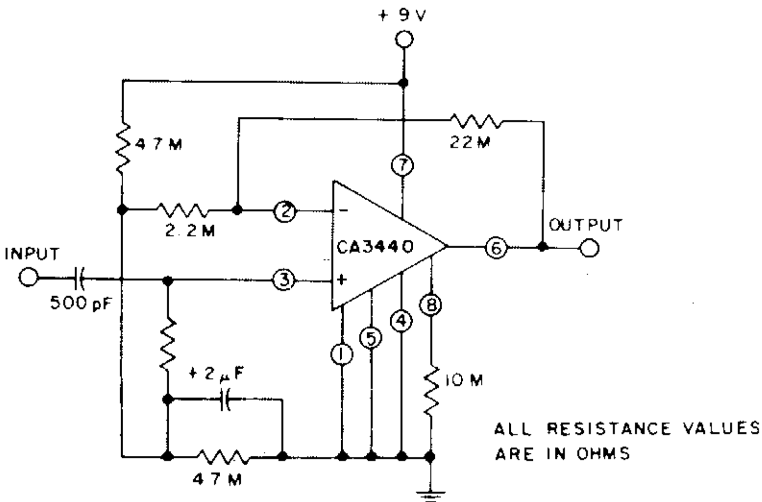


Fig. 6-4

R_{in} > 20 M
STAND-BY POWER = 90μW   3 kHz UPPER 3 dB BANDWIDTH
GAIN = 20 db
BW: 20-Hz TO 3-KHz
SR = 0.016 V/μs

GENERAL ELECTRIC/RCA

Circuit Notes

This circuit takes advantage of low power drain, high input impedance, and the excellent frequency capability of the CA3440. Only a 500-pF input coupling capacitor is needed to achieve a 20 Hz, −3 dB low-frequency response.

STEREO PREAMPLIFIER

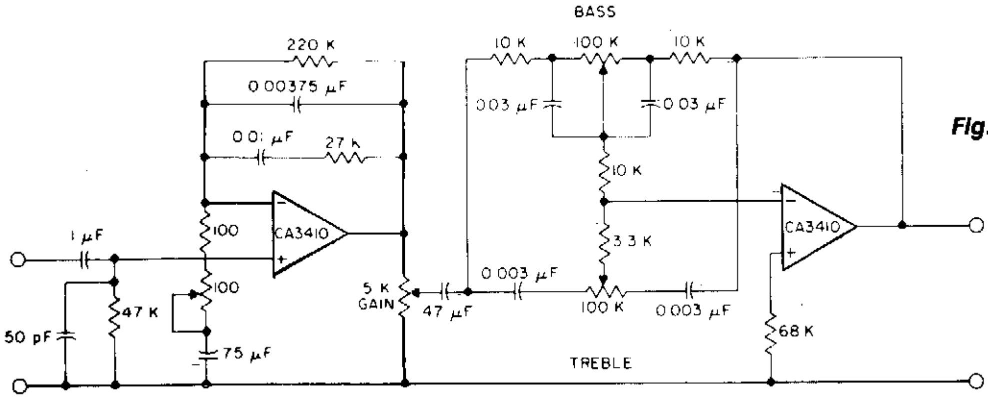


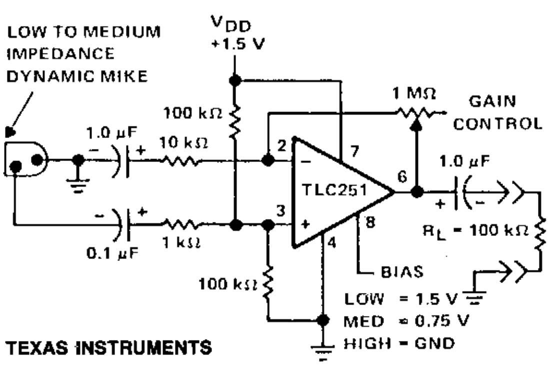
Fig. 6-5

GENERAL ELECTRIC/RCA

Circuit Notes

This circuit has RIAA playback equalization, tone controls, and adequate gain to drive a majority of commercial power amplifiers, using the CA3410 BiMOS op amp. Total harmonic distortion, when driven to provide a 6-V output, is less than 0.035% in the audio-frequency range of 150 Hz to 40 kHz. Complete stereo preamplifier consists of duplicating this circuit using the two remaining CA3410 amplifiers.

MICROPHONE PREAMPLIFIER



TEXAS INSTRUMENTS

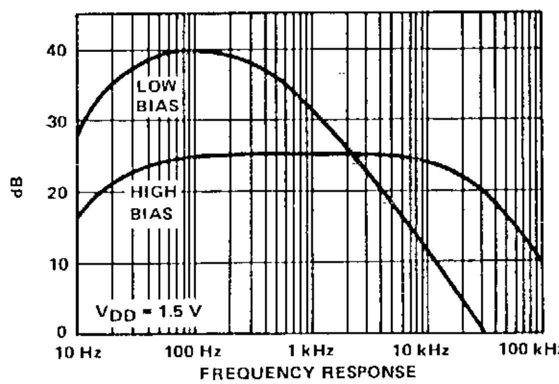
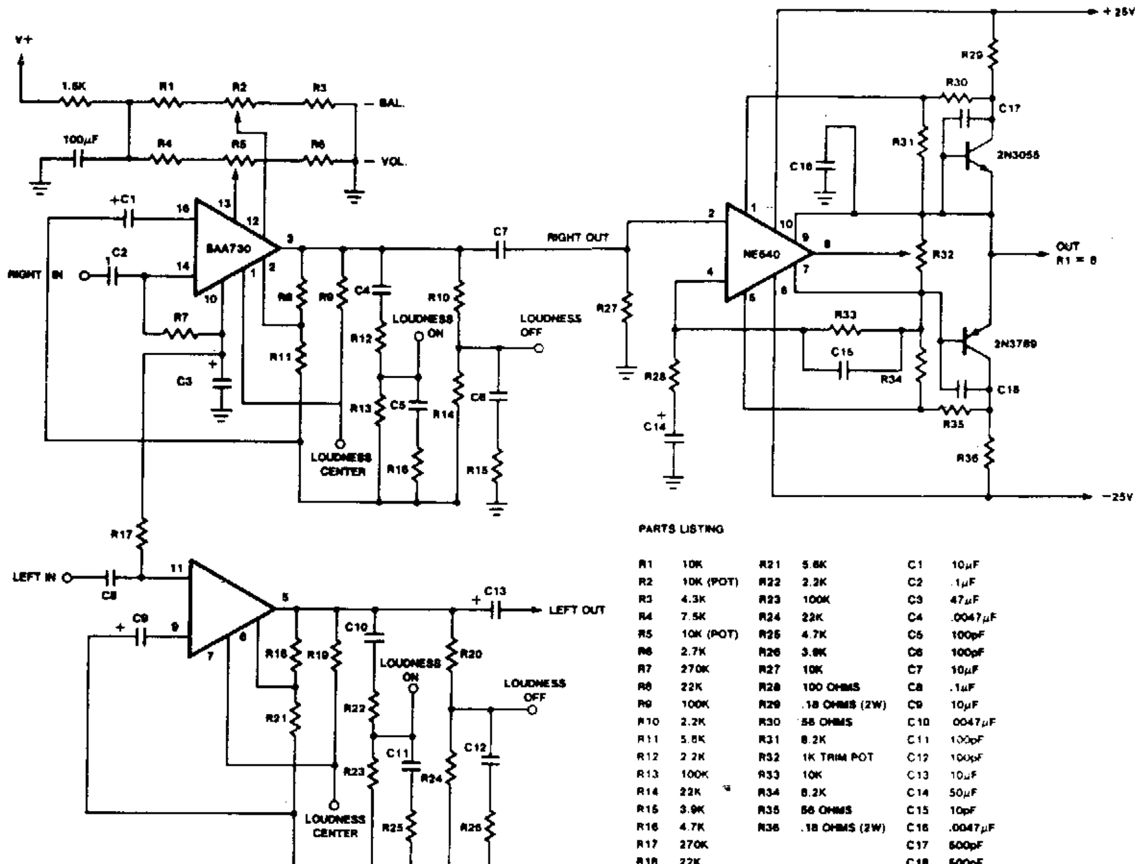


Fig. 6-6

Circuit Notes

A microphone preamplifier using a :om CMOS op amp complete with its own battery, is small enough to be put in a small mike case. The amplifier operates from a 1.5-V mercury cell battery at low supply currents. This preamplifier will operate at very low power levels and maintain a reasonable frequency response as well. The TLC251 operated in the low bias mode (operating at 1.5 V) draws a supply current of only 10 μA and has a −3 dB frequency response of 27 Hz to 4.8 kHz. With pin 8 grounded, which is designated as the high bias condition, the upper limit increases to 25 kHz. Supply current is only 30 μA under those conditions.

VOLUME, BALANCE, LOUDNESS & POWER AMPS



PARTS LISTING

R1	10K	R21	5.6K	C1	10μF
R2	10K (POT)	R22	2.2K	C2	.1μF
R3	4.3K	R23	100K	C3	47μF
R4	7.5K	R24	22K	C4	.0047μF
R5	10K (POT)	R25	4.7K	C5	100pF
R6	2.7K	R26	3.9K	C6	100pF
R7	270K	R27	10K	C7	10μF
R8	22K	R28	100 OHMS	C8	.1μF
R9	100K	R29	.18 OHMS (2W)	C9	10μF
R10	2.2K	R30	56 OHMS	C10	.0047μF
R11	5.6K	R31	8.2K	C11	100pF
R12	2.2K	R32	1K TRIM POT	C12	100pF
R13	100K	R33	10K	C13	10μF
R14	22K	R34	8.2K	C14	50μF
R15	3.9K	R35	56 OHMS	C15	10pF
R16	4.7K	R36	.18 OHMS (2W)	C16	.0047μF
R17	270K			C17	500pF
R18	22K			C18	500pF
R19	100K				
R20	2.2K				

NOTE
All resistors ¼W 1% to 5% unless otherwise noted

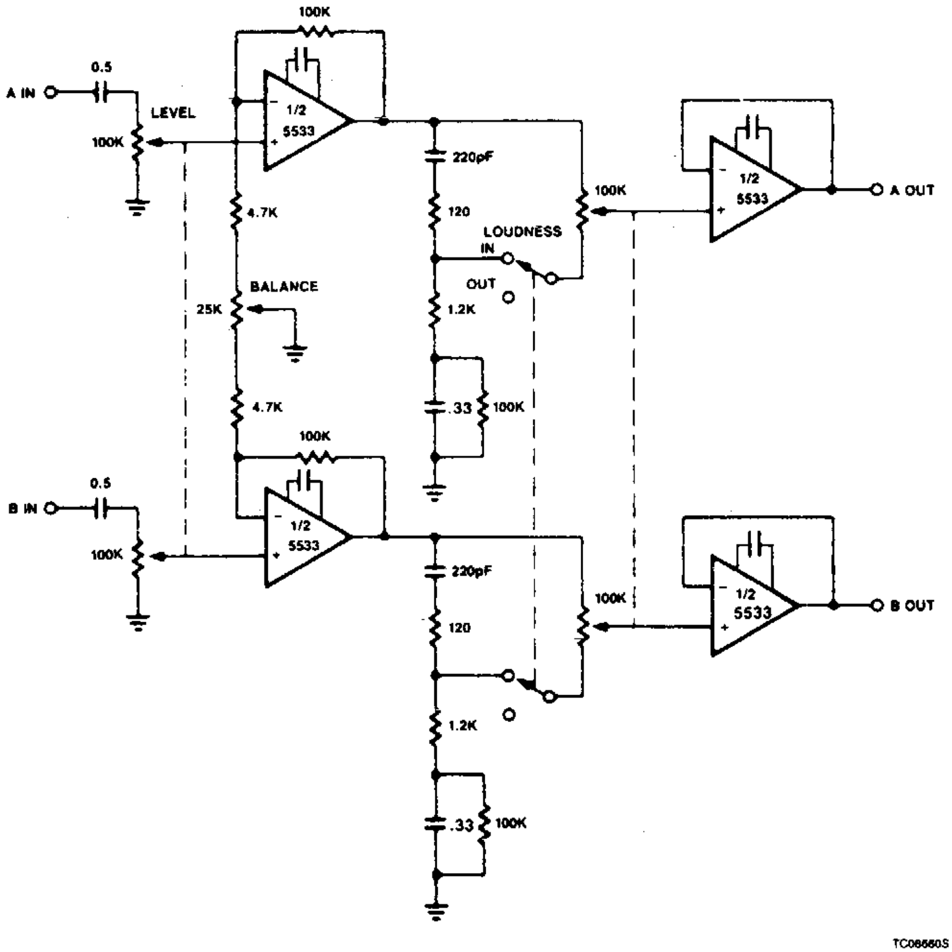
SIGNETICS

Fig. 6-7

Circuit Notes

This circuit should prove suitable as a design example for audio sound application.

BALANCE AND LOUDNESS AMPLIFIER



NOTE:
All resistor values are in ohms.

SIGNETICS *Fig. 6-7*

Circuit Notes

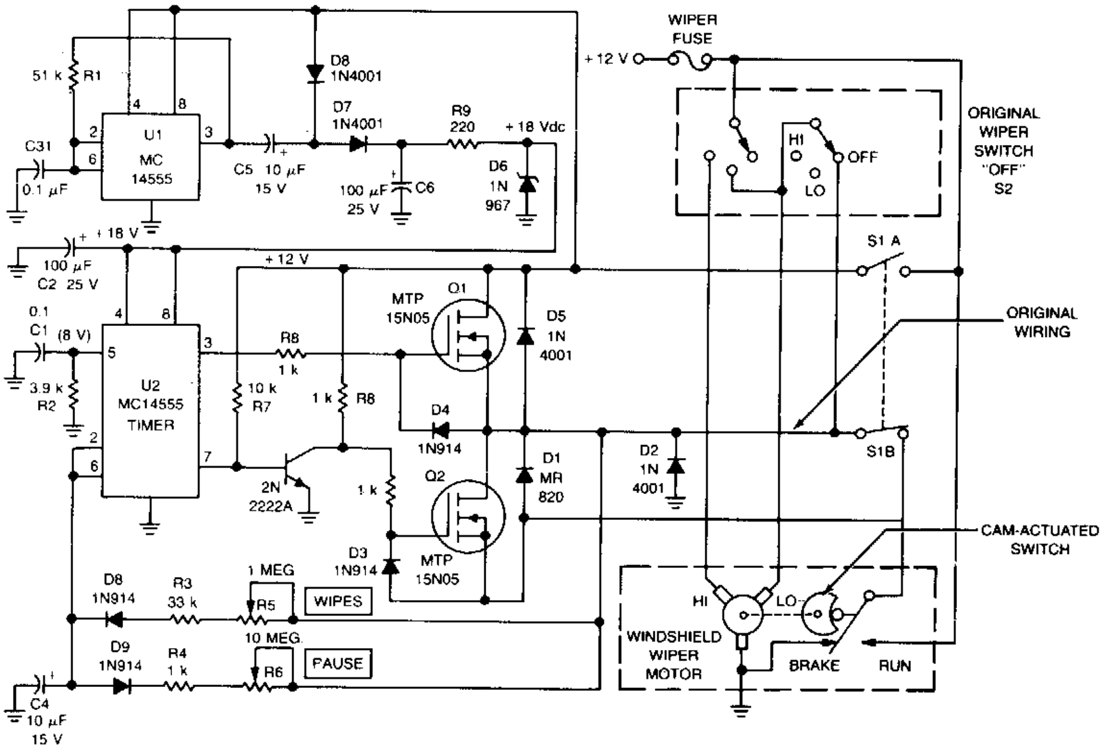
The circuit shows a combination of balance and loudness controls. Due to the non-linearity of the human hearing system, the low frequencies must be boosted at low listening levels. Balance, level, and loudness controls provide all the listening controls to produce the desired music response.

7

Automotive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

INTERMITTENT WINDSHIELD WIPER WITH DYNAMIC BRAKING



MOTOROLA

Fig. 7-1

Circuit Notes

The circuit provides a delayed windshield wiping, and dynamic braking of wiper blades when they reach the rest position. This prevents the blades from overshooting, which might cause them to stop at a point where they interfere with the drivers' vision.

With the original wiper switch off, switch S1A turns on the delay circuit and S1B disconnects the original automotive wiring. When S1 is turned off, the original wiring controls the system and the delay circuit is bypassed.

Turning S1 on applies the +12-V battery to U1 which is a voltage doubler that produces +18 V. This higher voltage supply is necessary to ensure reliable turn on of Q1 by multivibrator U2. This arrangement provides about +18 V to the gate of Q1, whose source is +12 V minus the V_{DS} drop of Q1.

Q1 remains on for a time determined by the WIPES potentiometer. The interval between wipes is controlled by the PAUSE control. When C1 drops below +4 V, U2 fires, turning Q1 on and restarting the cycle.

IMMOBILIZER

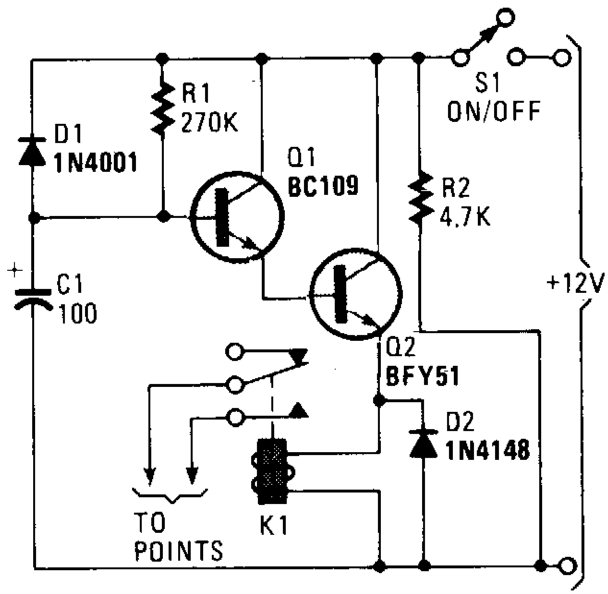


Fig. 7-2

HANDS-ON ELECTRONICS

Circuit Notes

A flip of S1 puts the circuit into action. Power for the circuit is picked up from the ignition switch, and the circuit receives no power until the ignition switch is closed. When power is turned on, capacitor C1 is not charged and the emitter-follower Darlington pair (formed by Q1 and Q2) are cutoff, thus no power is applied to the relay (K1), which serves as Q1's emitter load. The relay's normally-open contacts are connected across the vehicle's points. (At this time, the relay contacts are open and have no effect on the ignition system). C1 charges by way of R1, causing the voltage at the base of Q1 to rise steadily. That creates a similar rise in the voltage at the emitter of Q2. A Darlington pair is used to provide a high input-impedance, buffer stage so that the voltage across C2 is free to rise almost to the full supply potential. Loading effects do not limit the charge potential to just a few volts. Eventually, the voltage applied to the relay becomes sufficient to activate it. The contacts close and short out the points. The ignition system now doesn't act properly and the vehicle is disabled. If the ignition is switched off, power is removed from the circuit and diode D1, which was previously reverse-biased, is now forward biased by the charge on C1. D1 allows C1 to rapidly discharge through R2 (and any other dc paths across the supply lines). The circuit is ready to operate when the ignition is again turned on. The engine will operate, but not for very long. The values of R1 and C1 provides a delay of about 25 to 30 seconds. Increase R1's value to provide a longer delay.

AUTOMOTIVE EXHAUST EMISSIONS ANALYZER

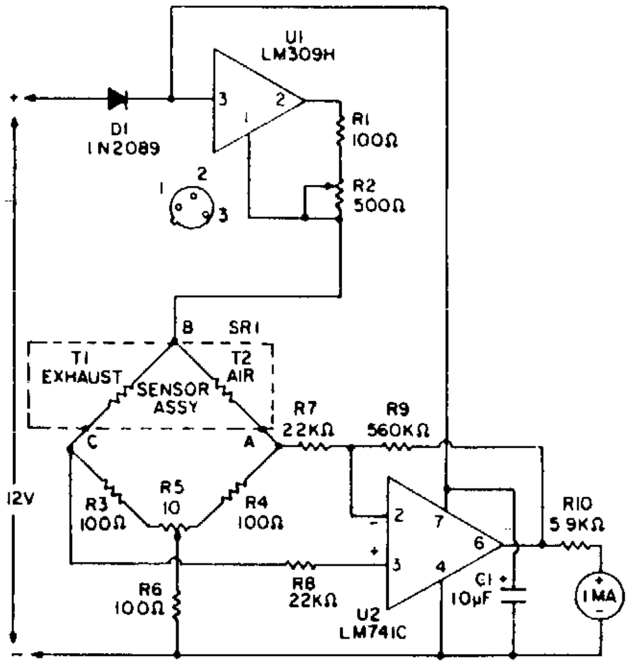


Fig. 7-3

C1—10-uF, 25 volt electrolytic capacitor
D1—Silicon diode, general purpose, 1N2069 or equivalent
M1—0.1 ma ammeter
R1, R3, R4, R6—100-ohm 10% ¼ watt resistor (All resistors are 10%, ¼ watt)
R2—500-ohm potentiometer, PC mount
R5—10-ohm potentiometer, front panel mount
R7, R8—22,000-ohm resistor
R9—570,000-ohm resistor
R10—5,900-ohm resistor
SR1—thermister sensor assembly, part number 100-1648 from Heathkit model CI-1080, Heath Company Benton Harbor, MI, 49022
U1—LM309H 5-volt regulator integrated circuit
U2—LM741TC op amp intergrated circuit

Misc.—PC Board, screws, cabinet, solder, hardware, 10-foot plastic tube, etc.

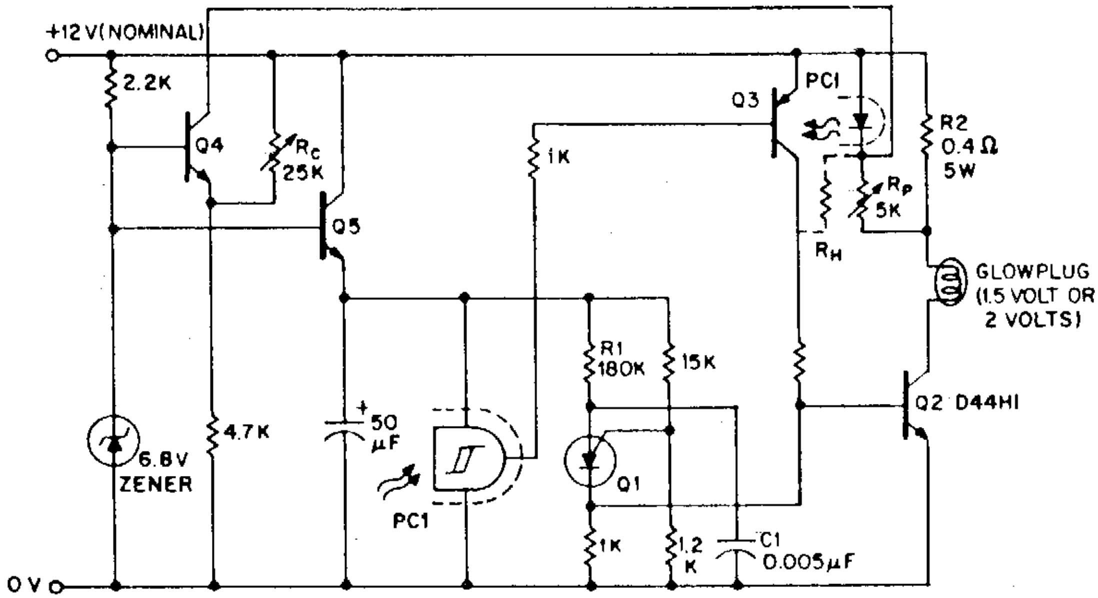
TAB BOOKS, INC.

Circuit Notes

A bridge circuit contains two 100-ohm resistors (R3 and R4), and two thermistors (T1 and T2). At room temperature the resistance of T1 and T2 is about 2000 ohms. When they are each heated to 150°C by a 10 mA current, the resistance value decreases to 100 ohms. Thus, the four elements comprise a bridge circuit. A characteristic of CO is that it conducts heat away from a thermistor at a different rate than air. One thermistor, T1, is exposed to the automobile exhaust while the other, T2, is isolated in a pure air environment. The difference in thermal conduction unbalances the bridge. A voltage difference is caused between points A and C. A differential amplifier, U1, amplifies this difference and drives the meter with sufficient current to read out the percentage of CO and the air-fuel ratio. A front panel balance control, R5, balances the bridge and calibrates the instrument. Calibration is performed when both thermistors are exposed to the outside air.

GLOW PLUG DRIVER

ALL RESISTORS 1/2W EXCEPT AS NOTED OTHERWISE



R_C — SUPPLY VOLTAGE COMPENSATION TRIM.
R_P — PLUG TEMPERATURE ADJUST.
Q1 — 2N6028
PC1 — H11L2.
Q3 — D29E2
Q4, Q5 — D33D1

GENERAL ELECTRIC

Fig. 7-4

Circuit Notes

Model airplanes, boats, and cars use glow plug ignitions for their miniature (0.8cc to 15cc) internal combustion engines. Such engines dispense with the heavy on-board batteries, H.T. coil, and "condenser" required for conventional spark ignition, while simultaneously developing much higher RPM (hence power) than the compression ignition (diesel) motors. The heart of a glow plug is a platinum alloy coil heated to incandescence for engine starting by an external battery, either 1.5 volts or 2 volts. Supplementing this battery, a second 12-volt power supply is frequently required for the engine starter, together with a third 6 volt type for the electrical fuel pump.

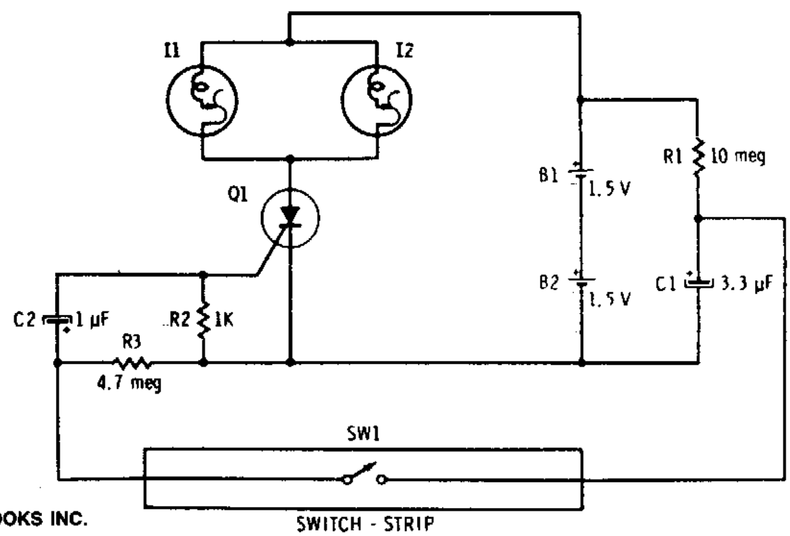
Rather than being burdened by all these multiple energy sources, the model builder would prefer to carry (and buy) a single 12-volt battery, deriving the lower voltages from this by use of suitable electronic step-down transformers (choppers). The glow driver illustrated does this and offers the additional benefit of (through negative feedback) maintaining constant plug temperature independent of engine flooding, or battery voltage while the starter is cranking.

In this circuit, the PUT relaxation oscillator Q1 turns on the output chopper transistor Q2 at a fixed repetition rate determined by R1 and C1. Current then flows through the glow plug and the parallel combination of the current sense resistor R2 and the LED associated with the H11L Schmitt trigger. With the plug cold (low resistance), current is high, the H11L is biased "on", and Q3 conducts to sustain base drive to Q2. Once the plug has attained optimum operating temperature, which can be monitored by its ohmic resistance, the H11L is programmed (via R_p) to switch off, removing base drive from Q3 and Q2.

However, since the H11L senses glow plug current, not resistance, this is only valid if supply voltage is constant, which is not always the case. Transistor Q4 provides suitable compensation in this case; if battery voltage falls (during cold cranking, for instance), the collector current of Q4 rises, causing additional current to flow through the LED, thus delaying the switch-off point for a given plug current. The circuit holds plug temperature relatively constant, with the plug either completely dry or thoroughly "wet", over an input voltage range of 8 to 16 volts. A similar configuration can be employed to maintain constant temperature for a full size truck diesel glow plug (28-volts supply, 12-volts glow plug); in this case, since plug temperature excursions are not so great, a hysteresis expansion resistor R_H may be required.

Fig. 7-4 Continued

GARAGE STOP LIGHT



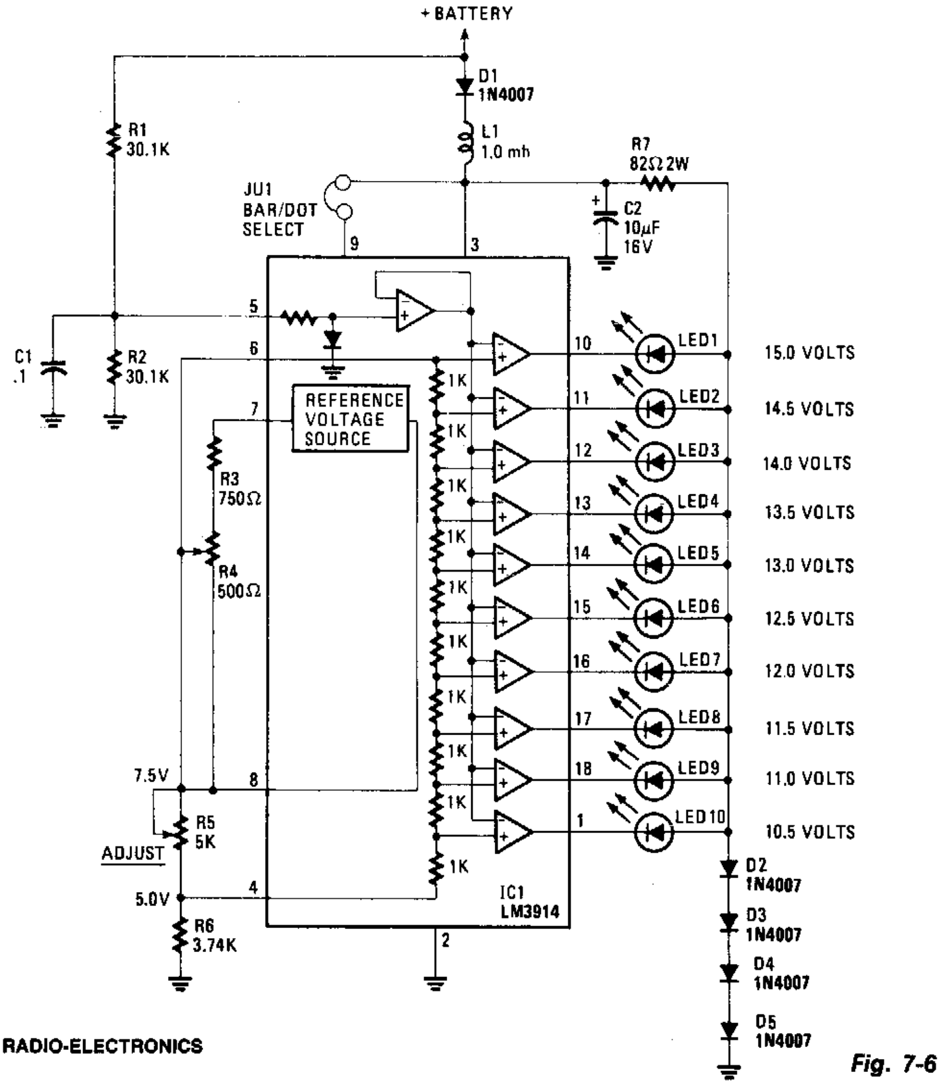
TAB BOOKS INC.

Fig. 7-5

Circuit Notes

Capacitor C1 is permanently connected across the 3-volt supply through 10 megohm resistor R1. The capacitor charges (relatively slowly) to 3 volts. The instant switch SW1 is closed, it connects the charged capacitor (C1) in series with C2 and R2. Capacitor C2 starts to charge, placing a positive-going voltage on the gate of the SCR and causing it to turn on. The two parallel-connected "self-flashing" bulbs I1 and I2 turn on. They flash and turn off the SCR and the circuit is off until car is driven off the switch and C1 can recharge.

BAR-GRAPH VOLTMETER



RADIO-ELECTRONICS

Fig. 7-6

Circuit Notes

This display uses ten LED's to display a voltage range from 10.5 to 15 volts. Each LED represents a 0.5-volt step in voltage. The heart of the circuit is the LM-3914 dot/bar display driver. Trimmer potentiometer R5 is adjusted so that 7.5 volts is applied to the top side of the divider. Resistor R7 and diodes D2 through D5 clamp the voltage applied to the LED's to about 3 volts. A lowpass filter made up of L1 and C2 guards against voltage spikes. Diode D1 is used to protect against reverse voltage in case the voltmeter is hooked up backward.

DELAYED-ACTION WINDSHIELD WIPER CONTROL

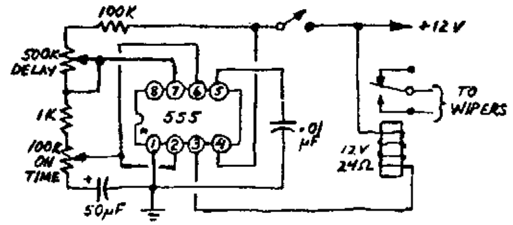
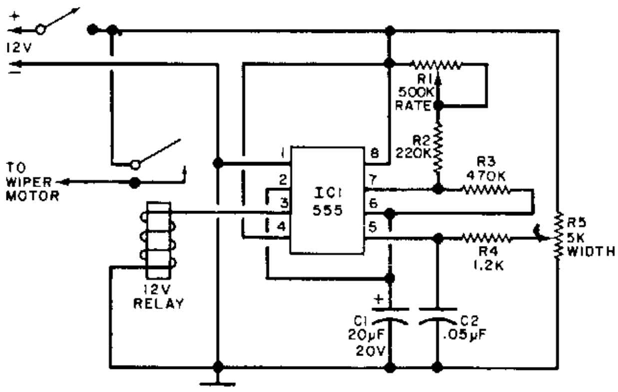


Fig. 7-7

POPULAR ELECTRONICS

SLOW-SWEEP WIPER CONTROL



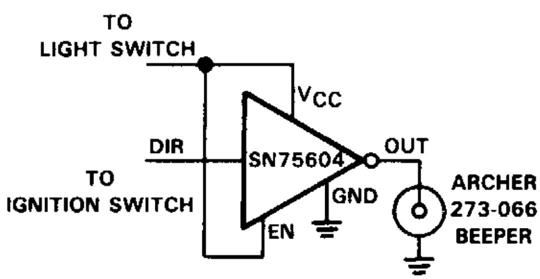
POPULAR ELECTRONICS

Circuit Notes

The relay which applies power to the wiper motor is actuated at periodic intervals by the timer circuit, closing the wiper motor contacts. Potentiometer R1 serves as the pulse rate control and potentiometer R5 as the pulse width control. These two controls should be adjusted for optimum performance after the unit is installed in a car.

Fig. 7-8

AUTOMOTIVE LIGHTS ON WARNING



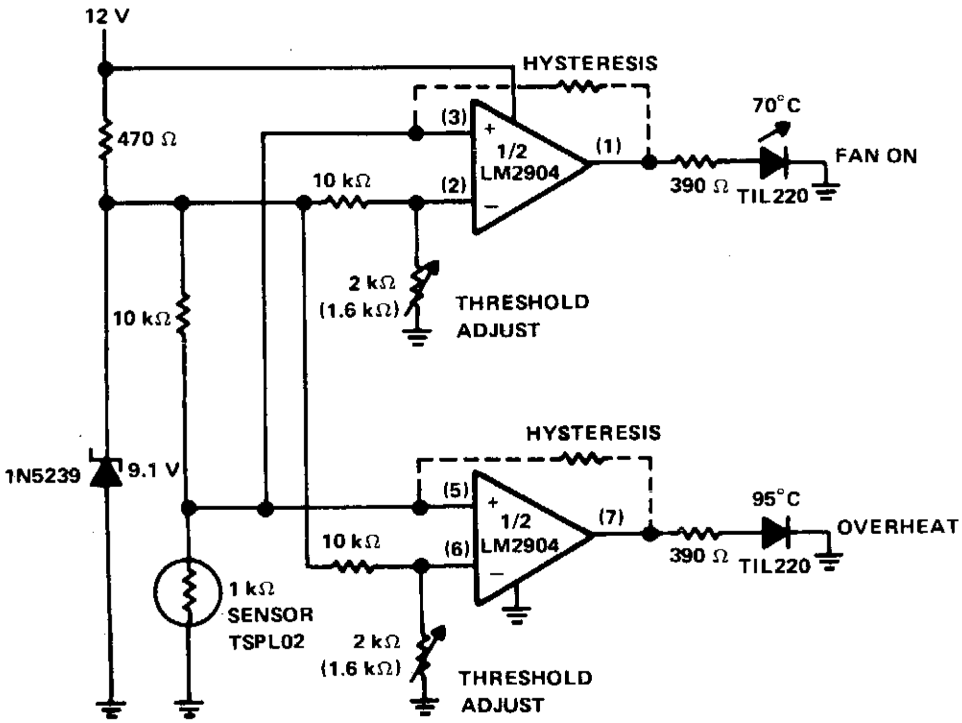
TEXAS INSTRUMENTS

Fig. 7-9

Circuit Notes

The SN75604, with input control logic but requiring only one supply rail, can be used in the "lights on" sensor and alarm driver. The device V_{CC} and enable inputs are connected to a voltage lead from the light switch. The direction control input is connected to a lead from the ignition switch. Only operation of the lights without the ignition will result in the alarm sounding. The beeper used in this application is an Archer 273-066 that will operate from 3 V to 28 V. At a typical 12 V level, it will produce a pulsating tone of about 95 dB at 30 cm. The alarm "on" current is about 12 mA when operating from a 12 V supply.

PTC THERMISTOR AUTOMOTIVE TEMPERATURE INDICATOR



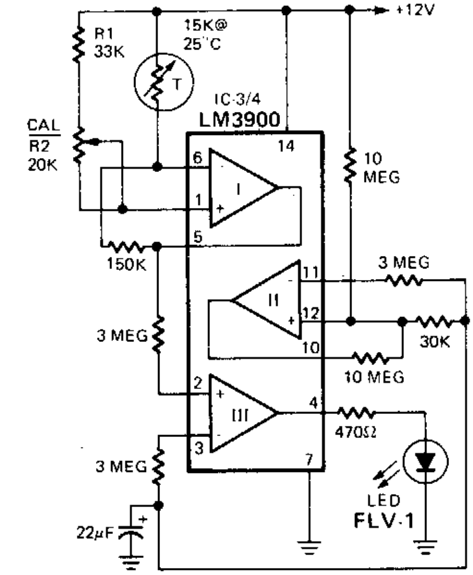
TEXAS INSTRUMENTS

Fig. 7-10

Circuit Notes

The circuit is used to indicate two different water temperature trip points by turning on LEDs when the temperatures are reached. The circuit is constructed around the LM2904 dual operational amplifier powered from the 12 V auto system. The thermistor is in series with a 10 kΩ resistor from ground to the positive 9.1 V point. The top of the thermistor is tied to both non-inverting inputs of the LM2904. The voltage at these inputs will change as the thermistor resistance changes with temperature. Each inverting input on the LM2904 has a reference, or threshold trip point, set by a 10 kΩ resistor and a 2 kΩ potentiometer in series across the 9.1 V regulated voltage. When this threshold is exceeded on the non-inverting input of LM2904, the TIL220 LED lights. The two trip points can be recalibrated or set to trip at different temperatures by adjusting the 2 kΩ potentiometer in each section. In addition to being used as warning lights as shown here, circuits can be added to turn on the fan motor or activate a relay.

ROAD ICE ALARM



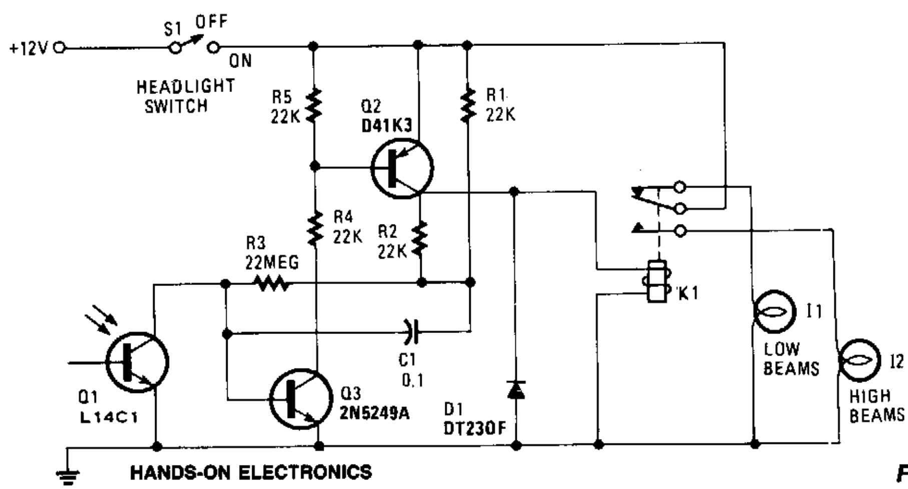
RADIO-ELECTRONICS

Fig. 7-11

Circuit Notes

The circuit uses a thermistor and three sections of a LM3900 quad op amp IC. When the temperature drops to 36°F the LED indicator flashes about once each second. The flashing rate increases as temperature drops to 32°F when the LED remains on. Amplifier I compares the thermistor's resistance to the resistance of the standard network connected to its noninverting input. Its output—fed to the noninverting input of op amp III—varies with temperature. Op amp II is a free-running multivibrator feeding a pulse signal of about 1 Hz to the inverting input of op amp III. This amplifier compares the outputs of op amps I and II and turns on the LED when the multivibrator's output level drops below op amp I. The monitor is calibrated by placing the thermistor in a mixture of crushed ice and water and adjusting the 20 kΩ pot so the LED stays on.

HEADLIGHT DIMMER



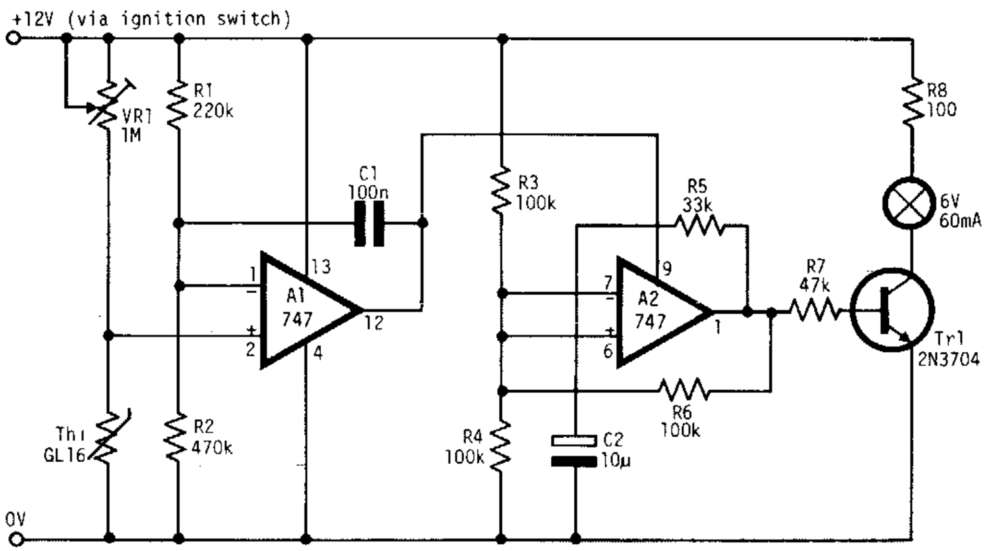
HANDS-ON ELECTRONICS

Fig. 7-12

Circuit Notes

When the lights of an on-coming car are sensed by photo-transistor Q1, things get going. Sensitivity is set by the 22-megohm resistor, R5, to about half a foot-candle. The relay used has a 12-volt, 0.3A coil. The L14C1 is complete with a lens that has a diameter of one inch for a 10° viewing angle.

ICE FORMATION ALARM



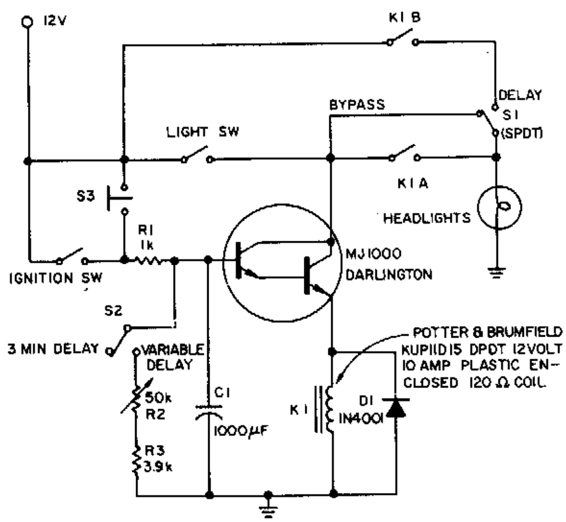
ELECTRONIC ENGINEERING

Fig. 7-13

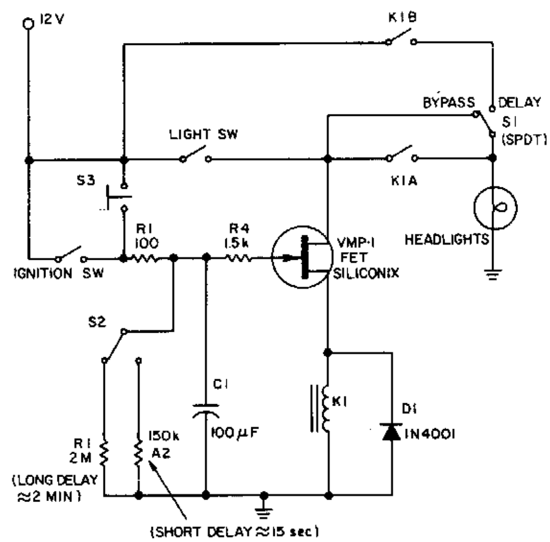
Circuit Notes

The circuit warns car drivers when the air temperature close to the ground approaches 0°C, thereby indicating possible formation of ice on the road surface. Op amp A1 is wired as a voltage level sensor. Op amp A2 is wired as an astable multivibrator which, by means of current buffer Tr1, flashes a filament lamp at about 1 Hz. As air temperature falls, a point is reached when the voltage at pin 2 just rises above the voltage at pin 1. The output of A1 is immediately driven into positive saturation, since it is operated open-loop. This positive output voltage powers A2 through its V + connection on pin 9, starting the oscillator. The thermistor is a glass bead type with a resistance of about 20 MΩ at 20°C. VR1 is adjusted so that the lamp starts flashing when the air temperature is 1 to 2°C.

DELAY CIRCUITS FOR HEADLIGHTS



1. **Automobile headlights may be kept on** up to 3 minutes after you leave the car with this Darlington time-delay circuit.



2. **A FET version of the delay circuit** allows the use of a smaller timing capacitor, C_1, for a given delay, and almost instantaneous reset with S_3; the Darlington circuit needs almost 2 s.

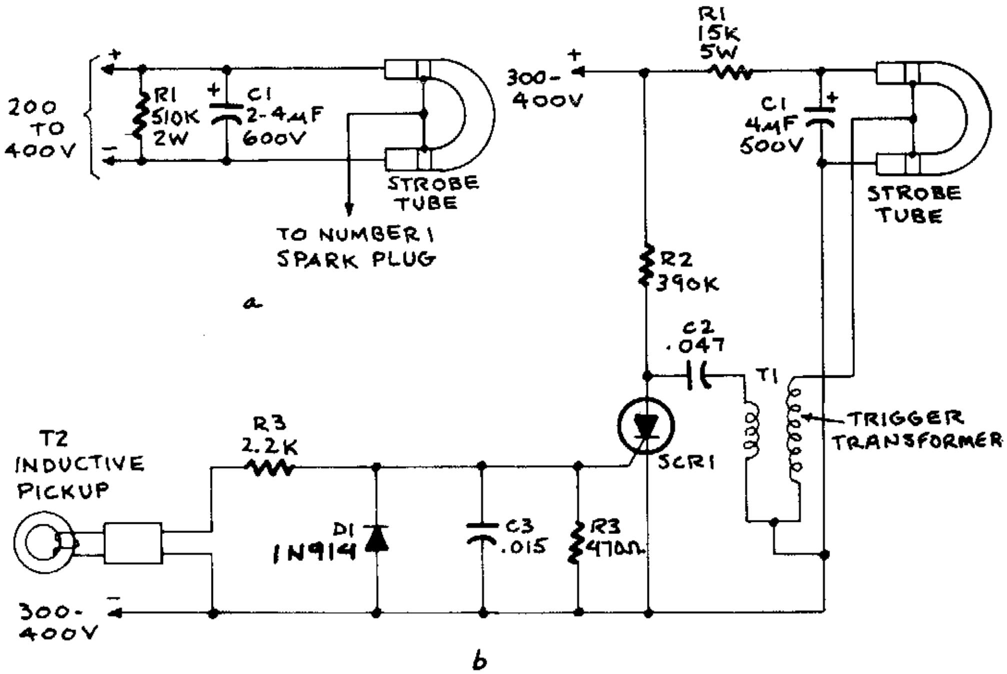
ELECTRONIC DESIGN

Fig. 7-14

Circuit Notes

This circuit keeps an automobile's headlights on temporarily. It also will turn the lights off, even if you forget to flip the light switch. The circuit's shut-off delay is actuated only after both the ignition and light switches have been on, and only if the ignition switch is turned off first. If the light switch is turned off first, no delay results. Parking and brake-light operation is not affected. The maximum time out can be up to 3 minutes in part 1 and hours with the circuit in part 2, depending on the relay selected and the value of R2. A switch S2 can be used to permit selection of either a short or long delay. Momentary switch S3 can restart circuit timing before the time-out is completed. A bypass switch, S1 removes the delay action.

IGNITION TIMING LIGHT



RADIO-ELECTRONICS

Fig. 7-15

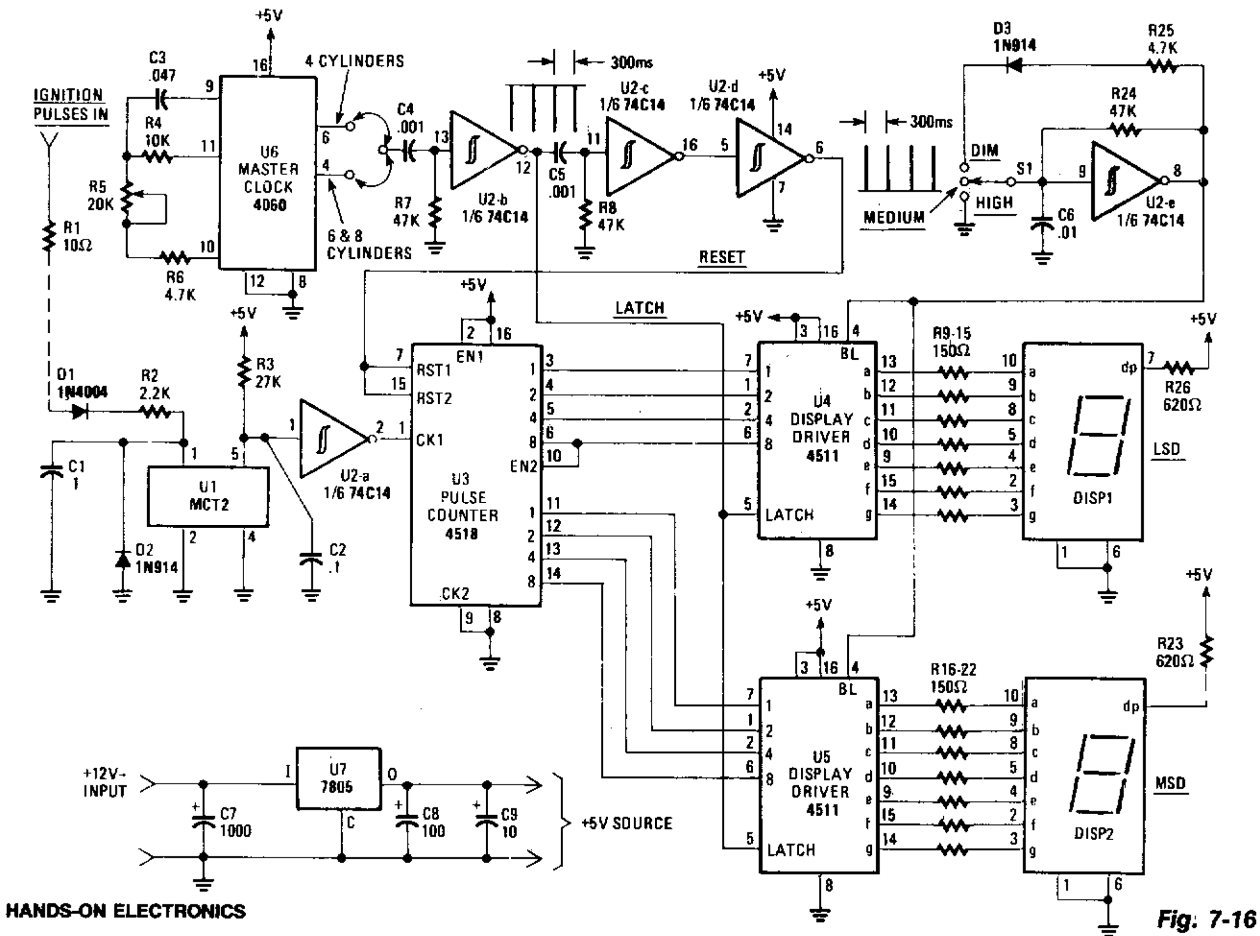
Circuit Notes

Figure *A* shows the circuit of a direct-trigger timing light. The trigger voltage is taken from the car's ignition circuit by a direct connection to a spark plug. A circuit using an inductive pickup is shown in Fig. *B*. A trigger transformer is used to develop the high-voltage pulse for triggering. The triggering circuit consists of T1, C1, SCR1, inductive pickup coil T2, and the waveshaping components in the SCR's gate circuit.

When the spark plug fires, it induces a pulse in pickup coil T2 that triggers the SCR gate. The SCR fires and discharges C2 through the primary of T1. The secondary of T1 feeds a high-voltage pulse to the trigger electrode of the flash tube. That pulse causes the gas—usually neon or xenon—to ionize. The ionized gas provides a low-resistance path for C1 to discharge, thereby creating a brilliant flash of light.

Resistor R1 limits current from the supply as the tube fires. When C1 is fully discharged the strobe tube cuts off and returns to its "high-resistance" state. The current through R2 is not enough to sustain conduction through SCR1, so it cuts off and remains off until it is re-triggered by a gate pulse.

DIGI-TACH



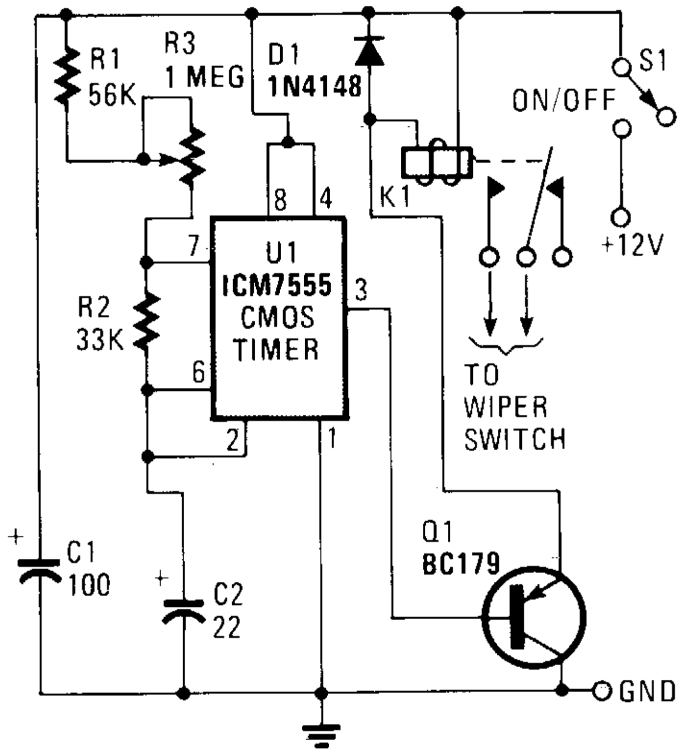
HANDS-ON ELECTRONICS

Fig. 7-16

Circuit Notes

The Digi-Tach contains a master-clock circuits (U6), latch and reset pulse generators (U2-b-U2-d), input signal conditioner (U1, U2-a), pulse counter (U3), display and display drivers (DIS1, DIS2, U4, and U5), and a voltage regulator (U7). As an added feature, Digi-Tach contains a dimmer circuit (U2-e).

CAR-WIPER CONTROL

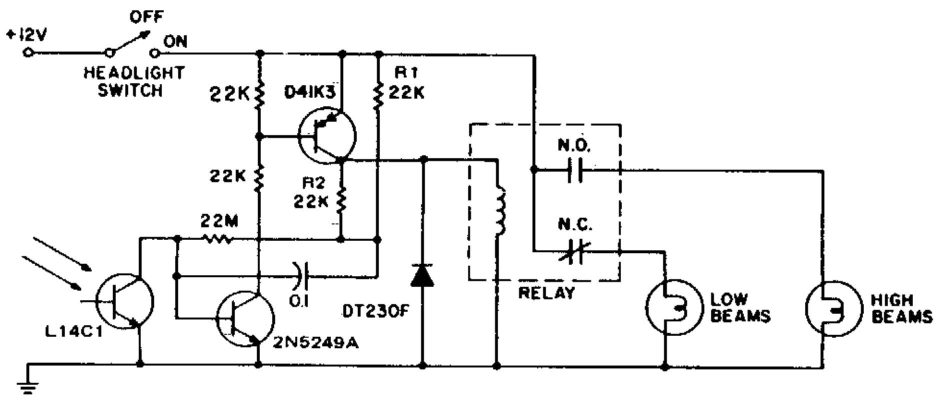


HANDS-ON ELECTRONICS *Fig. 7-17*

Circuit Notes

U1 is configured to operate in the standard astable mode, providing a form of relaxation oscillator. When power is applied, C2 initially charges through R1, R2 and R3 to two-thirds of the supply voltage. At that point, U1 senses that its threshold voltage at pin 6 has been reached, and triggers the timer, causing its output at pin 3 to go high. That high, applied to the base of Q1, keeps the transistor in the off state. Now C2 begins to discharge through R2 to pin 7 of U1. When C2 has discharged to about one-third of the supply voltage, U1 is toggled back to its original state. C2 starts to charge again, as pin 3 of U1 goes low. The low at pin 3 causes Q1—which serves as an emitter-follower buffer stage—to turn on, allowing current to flow through the coil of relay K1. That, in turn, causes K1's contacts to close, applying power to the wipers. The charge time of capacitor C2 is determined by the setting of potentiometer R3. Capacitor C2 should be a tantalum type, and actually, almost any 12-volt coil relay with sufficiently heavy contacts should serve well.

AUTOMATIC HEADLIGHT DIMMER



RELAY: 12V, 0.3A COIL: 20A, FORM C, CONTACTS OR SOLID-STATE SWITCHING OF 16A STEADY-STATE 150A COLD FILAMENT SURGE, RATING.

LENS: MINIMUM 1" DIAMETER, POSITIONED FOR ABOUT 10° VIEW ANGLE.

GENERAL ELECTRIC

Fig. 7-18

Circuit Notes

This circuit switches car headlights to the low beam state when it senses the lights of an on-coming car. The received light is very low level and highly directional, indicating the use of a lens with the detector. A relatively large amount of hysteresis is built into the circuit to prevent "flashing lights." Sensitivity is set by the 22 megohm resistor to about 0.5 ft. candle at the transistor (0.01 at the lens), while hysteresis is determined by the R1, R2 resistor voltage divider, parallel to the D41K3 collector emitter, which drives the 22 megohm resistor; maximum switching rate is limited by the 0.1 μF capacitor to 15/minute.

8

Battery Chargers and Zappers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Rapid Battery Charger for ICOM IC-2A
Gel Cell Charger
Ni-Cad Battery Zapper
Lithium Battery Charger
Thermally Controlled Ni-Cad Charger
Ni-Cad Battery Zapper II
Battery Charger
Wind Powered Battery Charger
Battery Charger Operates On Single Solar Cell
Versatile Battery Charger
14-Volt, 4-Amp Battery Charger/Power Supply

RAPID BATTERY CHARGER FOR ICOM IC-2A

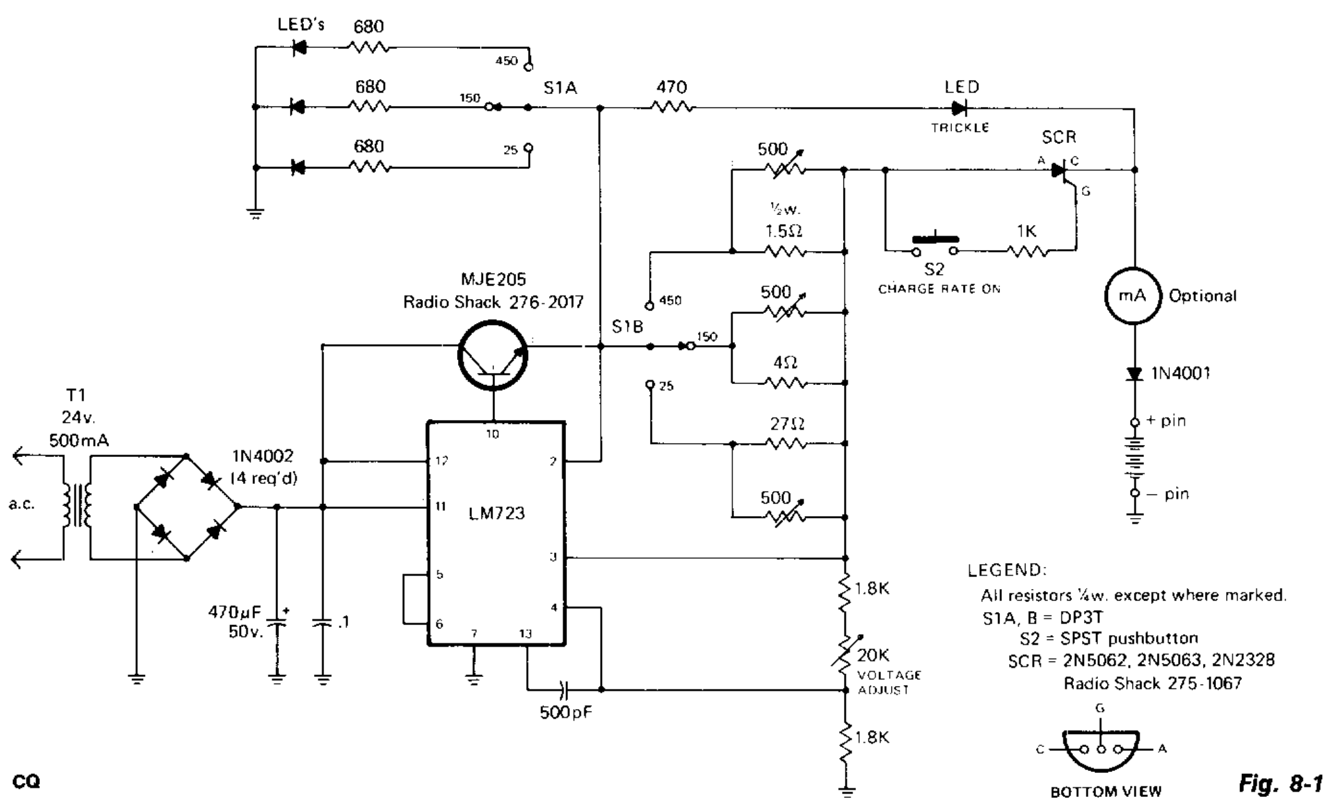
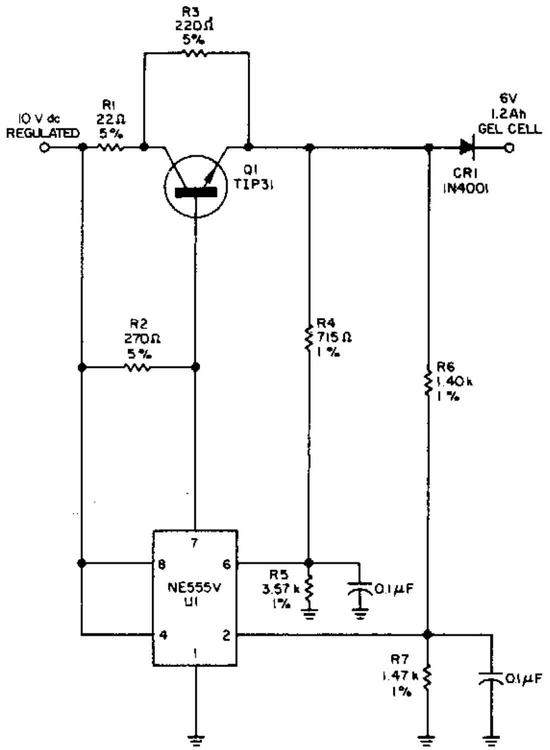


Fig. 8-1

Circuit Notes

Rectified and filtered voltage from the 24 Vac transformer is applied to the LM723 voltage regulator and the npn pass transistor set up for constant current supply. The 470 ohm resistor limits trickle current until the momentary pushbutton (S2) is depressed, the SCR turns on and current flows through the previously determined resistor network limiting the charging current. The SCR will turn off when the thermal cutout circuit inside the battery pack opens up.

GEL CELL CHARGER



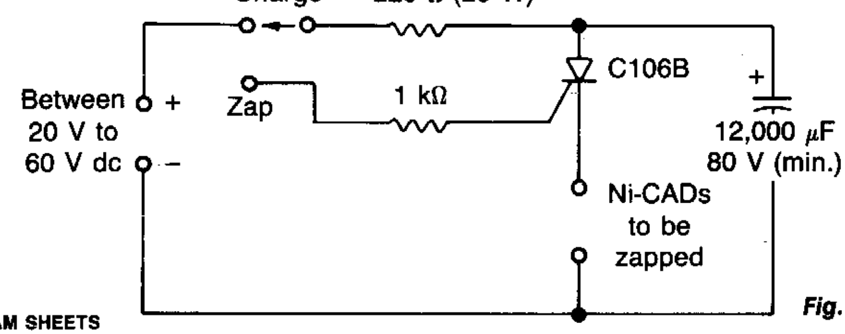
ELECTRONIC DESIGN

Circuit Notes

This circuit detects a full-charge state and automatically switches to a float condition—from 240 mA to 12 mA.

Fig. 8-2

Ni-CAD BATTERY ZAPPER



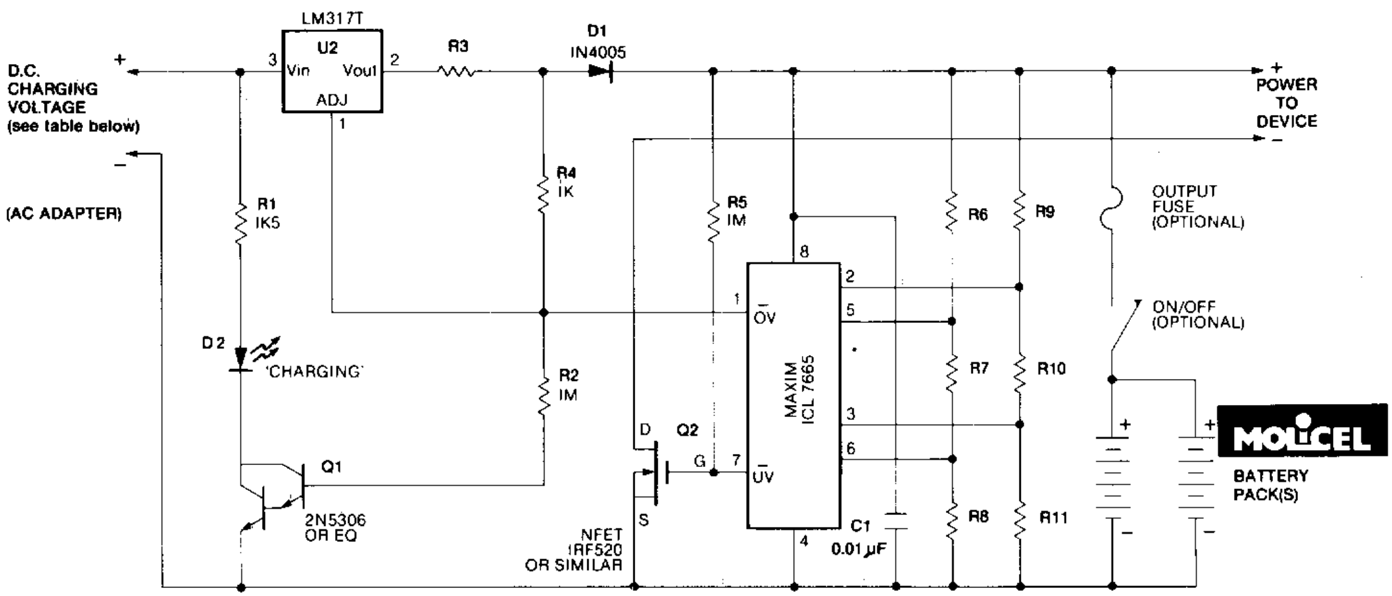
WILLIAM SHEETS

Fig. 8-3

Circuit Notes

The short in a Ni-Cad battery can be "burned off" with this zapper. Use of the SCR keeps heavy discharge current from damaging switch contacts.

LITHIUM BATTERY CHARGER



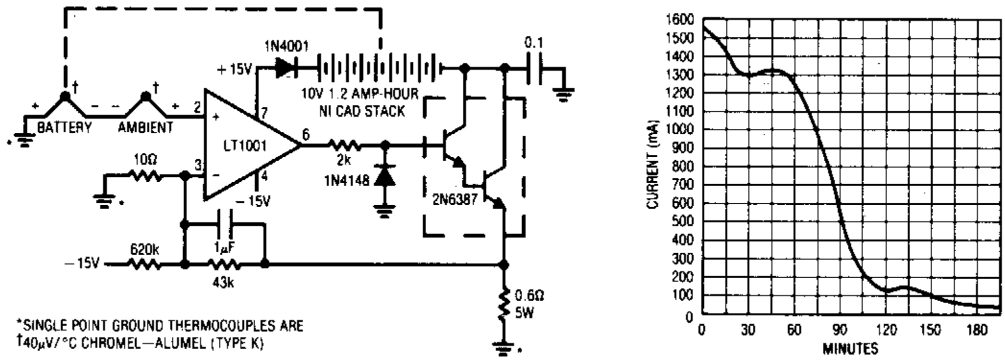
MOLI ENERGY LIMITED

Fig. 8-4

Circuit Notes

Charging is accomplished with a constant current of 60 mA for AA cells to a cutoff voltage of 2.4 V per cell at which point the charge must be terminated. The charging system shown is designed for multi-cell battery packs of 2 to 6 series-connected cells or series/parallel arrangements. It is essential that all cells assembled in the pack be at an identical state-of-charge (voltage) prior to charging. The maximum upper cut-off voltage is 15.6 volts (6×2.6 V).

THERMALLY CONTROLLED Ni-CAD CHARGER

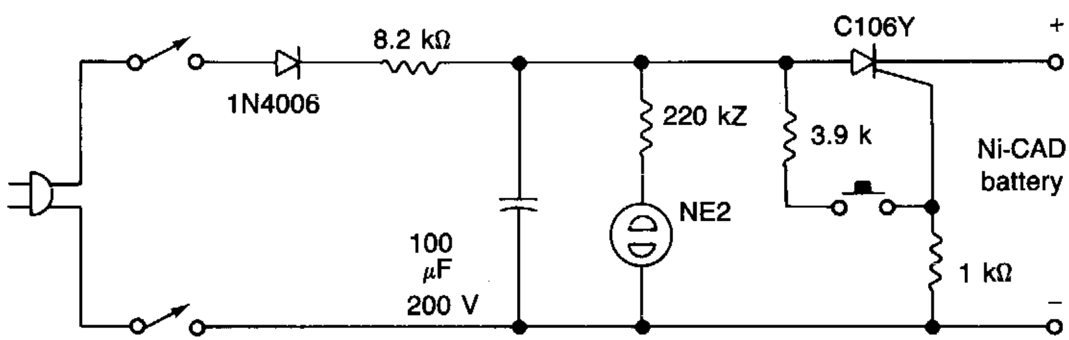


LINEAR TECHNOLOGY CORPORATION ***Fig. 8-5***

Circuit Notes

One way to charge Ni-Cad batteries rapidly without abuse is to measure cell temperature and taper the charge accordingly. The circuit uses a thermocouple for this function. A second thermocouple nulls out the effects of ambient temperature. The temperature difference between the two thermocouples determines the voltage which appears at the amplifier's positive input. As battery temperature rises, this small negative voltage (1°C difference between the thermocouples equals 40 μV) becomes larger. The amplifier, operating at a gain of 4300, gradually reduces the current through the battery to maintain its inputs at balance. The battery charges at a high rate until heating occurs and the circuit then tapers the charge. The values given in the circuit limit the battery surface temperature rise over ambient to about 5°C.

Ni-CAD BATTERY ZAPPER II

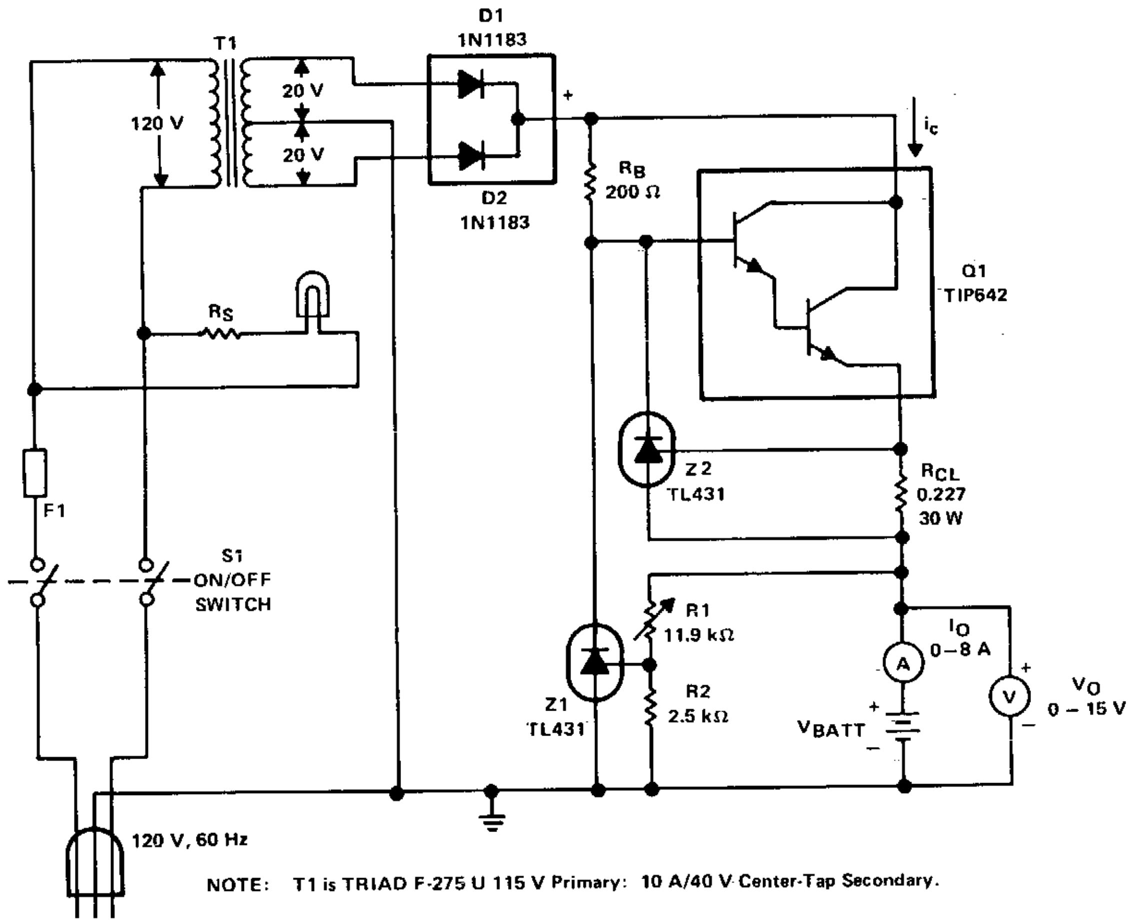


WILLIAM SHEETS ***Fig. 8-6***

Circuit Notes

This zapper clears internal short in nickel cadmium batteries by burning it away. CAUTION: The negative battery terminal is connected to one side of the ac line. For safety operation use a 1:1 isolation transformer.

BATTERY CHARGER



NOTE: T1 is TRIAD F-275 U 115 V Primary: 10 A/40 V Center-Tap Secondary.

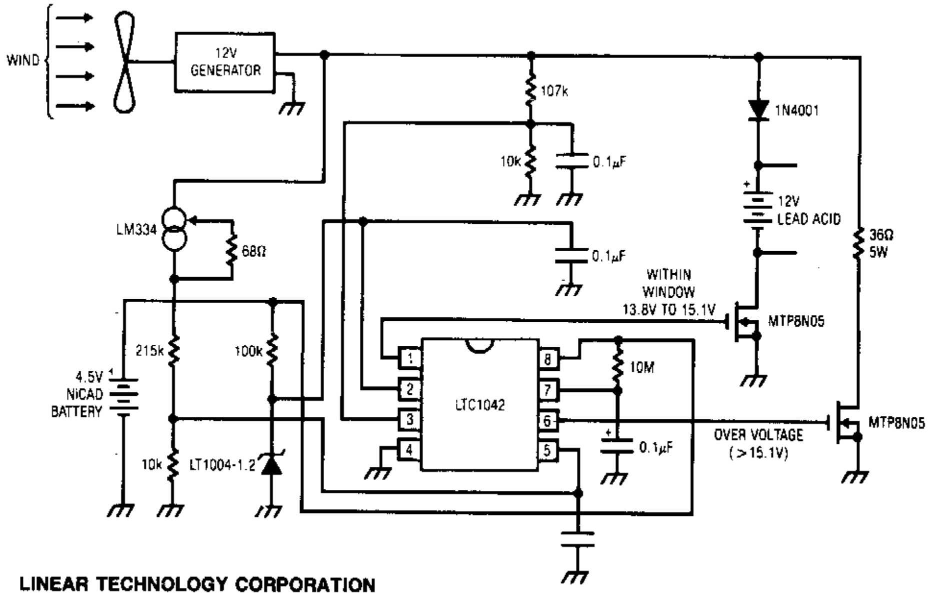
TEXAS INSTRUMENTS

Fig. 8-7

Circuit Notes

The charger is based on a charging voltage of 2.4 V per cell, in accordance with most manufacturers' recommendations. The circuit pulses the battery under charge with 14.4 V (6 cells × 2.4 V per cell) at a rate of 120 Hz. The design provides current limiting to protect the charger's internal components while limiting the charging rate to prevent damaging severely discharged lead-acid batteries. The maximum recommended charging current is normally about one-fourth the ampere-hour rating of the battery. For example, the maximum charging current for an average 44 ampere-hour battery is 11 A. If the impedance of the load requires a charging current greater than the 11 A current limit, the circuit will go into current limiting. The amplitude of the charging pulses is controlled to maintain a maximum peak charging current of 11 A (8 A average).

WIND POWERED BATTERY CHARGER



LINEAR TECHNOLOGY CORPORATION

Fig. 8-8

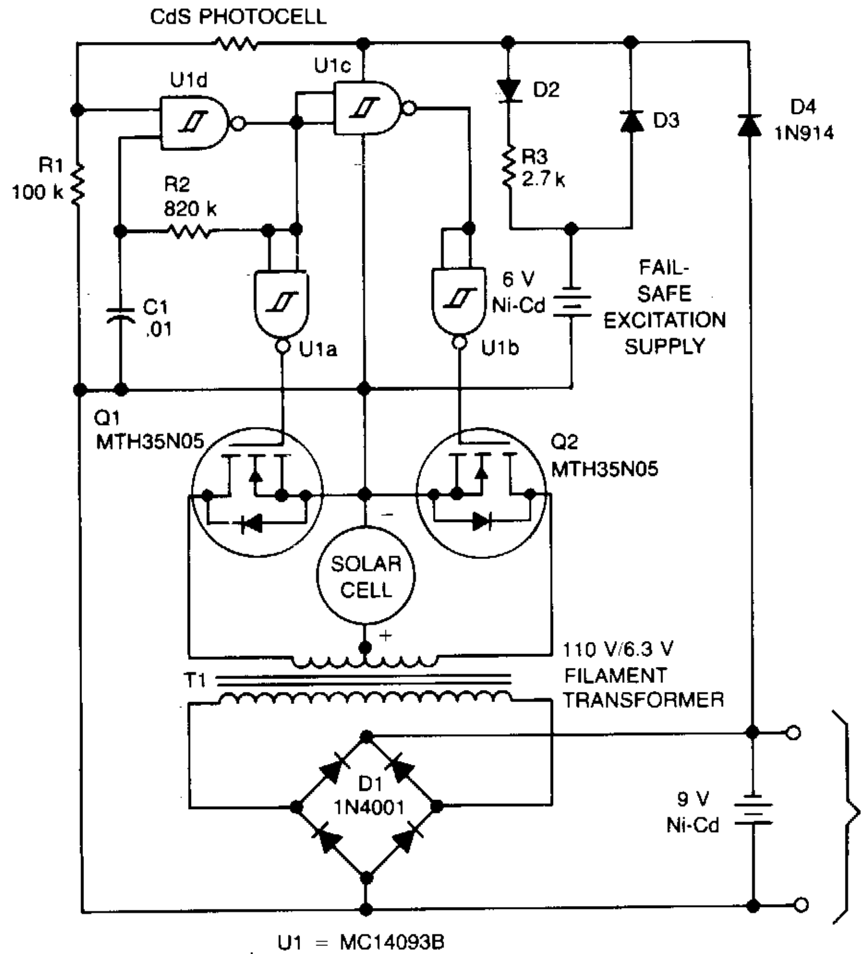
Circuit Notes

The dc motor is used as a generator with the voltage output being proportional to its rpm. The LTC1042 monitors the voltage output and provides the following control functions.

1. If generator voltage output is below 13.8 V, the control circuit is active and the Ni-Cad battery is charging through the LM334 current source. The lead acid battery is not being charged.
2. If the generator voltage output is between 13.8 V and 15.1 V, the 12 V lead acid battery is being charged at about 1 amp/hour rate (limited by the power FET).
3. If generator voltage exceeds 15.1 V (a condition caused by excessive wind speed or 12 V battery being fully charged) then a fixed load is connected limiting the generator rpm to prevent damage.

This charger can be used as a remote source of power where wind energy is plentiful such as on sailboats or remote radio repeater sites. Unlike solar powered panels, this system will function in bad weather and at night.

BATTERY CHARGER OPERATES ON SINGLE SOLAR CELL



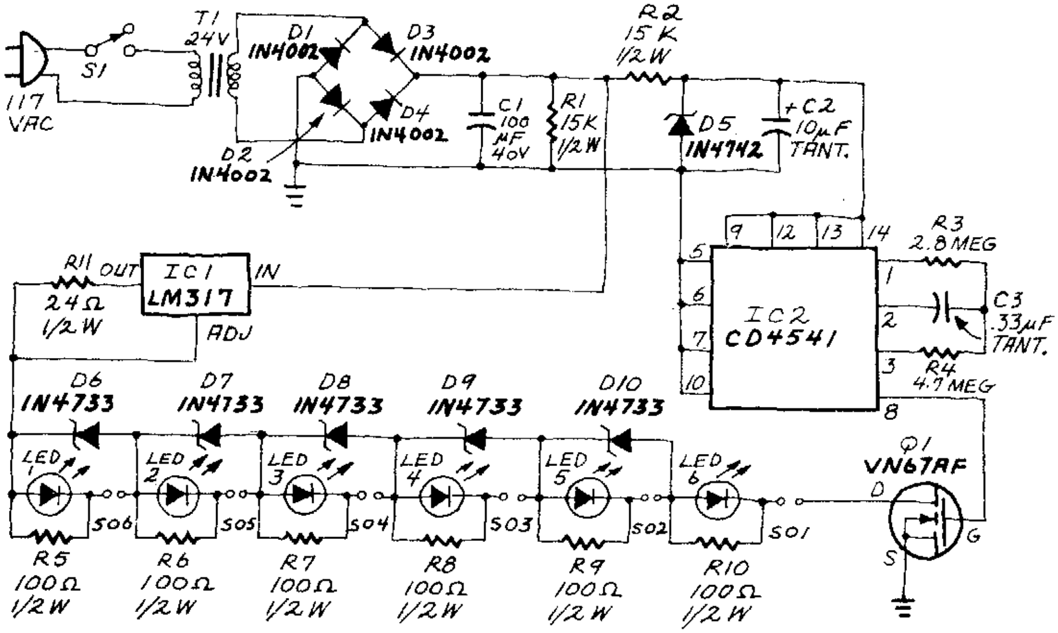
MOTOROLA *Fig. 8-9*

Circuit Notes

The circuit charges a 9-V battery at about 30 mA per input ampere at 0.4 V. U1, a quad Schmitt trigger, operate as an astable multivibrator to drive push-pull TMOS devices Q1 and Q2. Power for U1 is derived from the 9-V battery via D4; power for Q1 and Q2 is supplied by the solar cell. The multivibrator frequency, determined by R2-C1, is set to 180 Hz for maximum efficiency from a 6.3-V filament transformer, T1. The secondary of the transformer is applied to a full wave bridge rectifier, D1, which is connected to the batteries being charged. The small Ni-Cad battery is a fail-safe excitation supply to allow the system to recover if the 9-V battery becomes fully discharged.

A CdS photocell shuts off the oscillator in darkness to preserve the fail-safe battery during shipping and storage, or prolonged darkness.

VERSATILE BATTERY CHARGER



RADIO ELECTRONICS

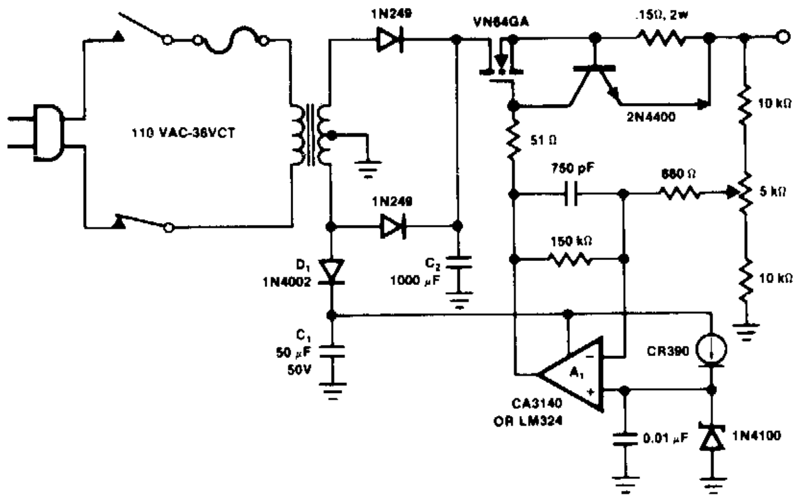
Fig. 8-10

Circuit Notes

An LM317 voltage regulator is configured as a constant-current source. It is used to supply the 50 mA charging current to S01-S06, an array of AA-cell battery holders. Each of the battery holders is wired in series with an LED and its associated shunt resistor. When the battery holder contains a battery, the LED glows during charging. Each battery holder/LED combination is paralleled by a 5.1-volt Zener diode. If the battery holder is empty, the Zener conducts the current around the holder.

A timing circuit prevents overcharging. When power is applied to the circuit, timing is initiated by IC2, a CD4541 oscillator/programmable timer. The output of IC2 is fed to Q1. When that output is high, the transistor is on, and the charging circuit is completed. When the output is low, the transistor is off, and the path to ground is interrupted.

14-VOLT, 4-AMP BATTERY CHARGER/POWER SUPPLY



SILICONIX, INC.

Fig. 8-11

Circuit Notes

Operation amplifier A1 directly drives the VN64GA with the error signal to control the output voltage. Peak rectifier D1, C1 supplies error amplifier A1 and the reference zener. This extra drive voltage must exceed its source voltage by several volts for the VN64GA to pass full load current. The output voltage is pulsating dc which is quite satisfactory for battery charging. To convert the system to a regulated dc supply, capacitor C2 is increased and another electrolytic capacitor is added across the load. The response time is very fast, being determined by the op-amp. The 2N4400 current limiter circuit prevents the output current from exceeding 4.5 A. However, maintaining a shorted condition for more than a second will cause the VN64GA to exceed its temperature ratings. A generous heat sink, on the order of 1°C/W, must be used.

9

Battery Monitors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DYNAMIC, CONSTANT CURRENT LOAD FOR FUEL CELL/BATTERY TESTING

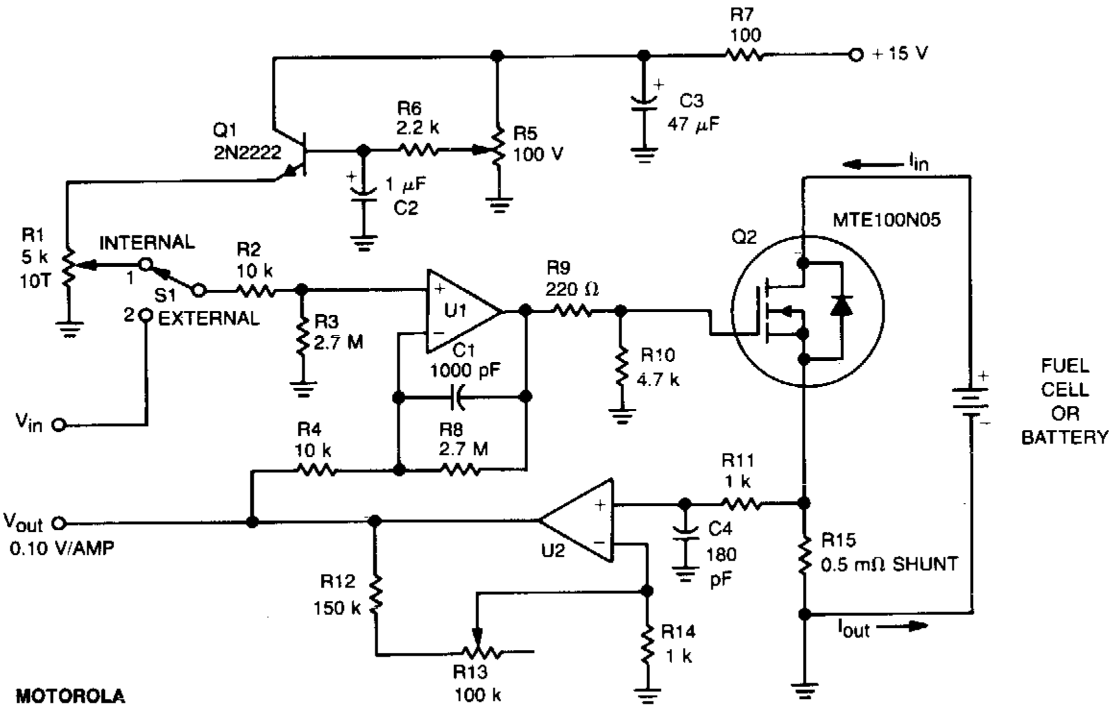


Fig. 9-1

Circuit Notes

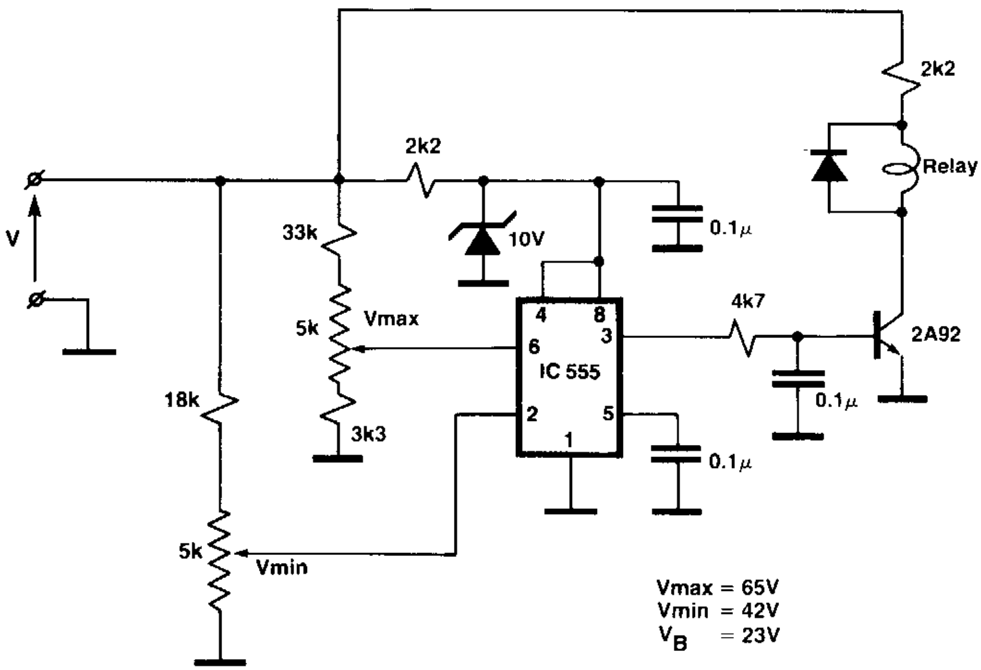
This circuit was designed for testing fuel cells, but it could also be used for testing batteries under a constant current load. It provides a dynamic, constant current load, eliminating the need to manually adjust the load to maintain a constant load.

For fuel cell application, the load must be able to absorb 20-40 A, and since a single cell develops only 0.5 to 1.0 V, bipolar power devices (such as a Darlington) are impractical. Therefore, this dynamic load was designed with a TMOS Power FET (Q2).

With switch S1 in position 1, emitter follower Q1 and R1 establish the current level for the load. In position 2, an external voltage can be applied to control the current level.

Operational amplifier U1 drives TMOS device Q1, which sets the load current seen by the fuel cell or battery. The voltage drop across R15, which is related to the load current, is then applied to U2, whose output is fed back to U1. Thus, if the voltage across R15 would tend to change, feedback to the minus input of U1 causes that voltage (and the load current) to remain constant. Adjustment of R13 controls the volts/amp of feedback. The V_{OUT} point is used to monitor the system.

VOLTAGE DETECTOR RELAY FOR BATTERY CHARGER



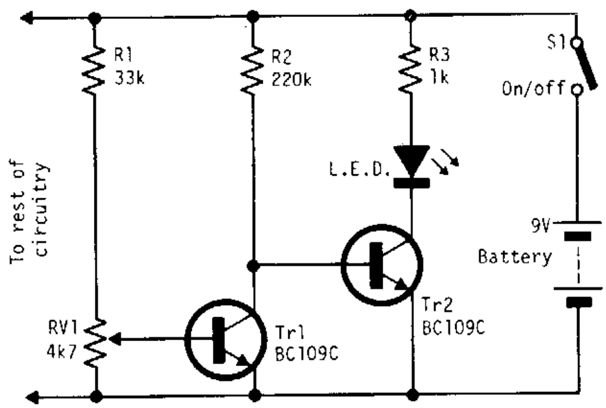
ELECTRONIC ENGINEERING

Fig. 9-2

Circuit Notes

While the battery is being charged, its voltage is measured at V. If the measured voltage is lower than the minimum the relay will be energized, that will connect the charger circuit. When the battery voltage runs over the maximum set point, the relay is deenergized and it will be held that way until the voltage decreases below the minimum when it will be connected again. The voltage is lower than a threshold V_B (low breaking voltage) the relay will be assumed that such a low voltage is due to one or several damaged battery components. Of course V_B is much lower than the minimum set point.

BATTERY STATUS INDICATOR



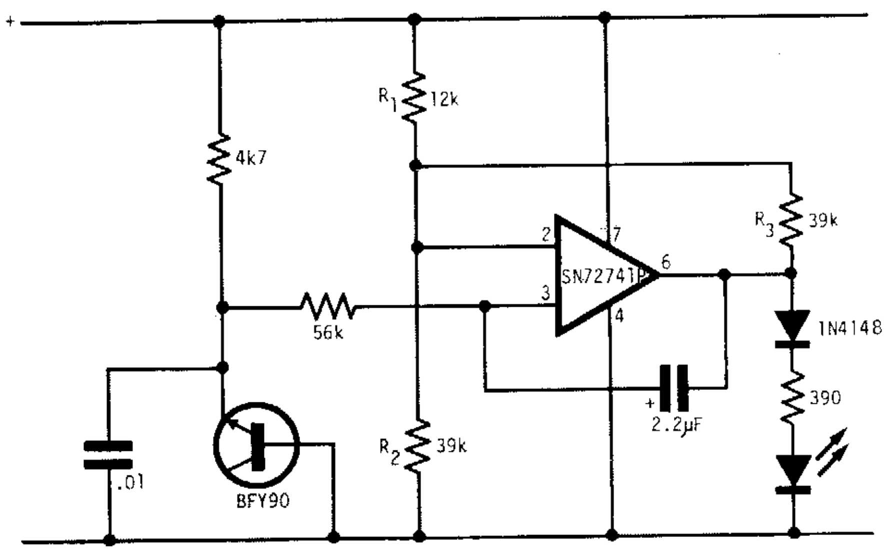
ELECTRONIC ENGINEERING

Fig. 9-3

Circuit Notes

Continually monitors battery voltage during use and consumes only about 250 μA (until the end point is reached). Near the end point Tr1 turns off, allowing Tr2 to illuminate the LED to increase current drain further leading to a distinct turn off point.

LOW-BATTERY INDICATOR



ELECTRONIC ENGINEERING

Fig. 9-4

Circuit Notes

Under good battery conditions the LED is off. As the battery voltage falls, the LED begins to flash until, in the low battery condition, the LED lights continuously. Designed for a 9-volt battery, with the values shown the LED flashes from 7.5 to 6.5 volts.

A LITHIUM BATTERY'S STATE-OF-CHARGE INDICATOR

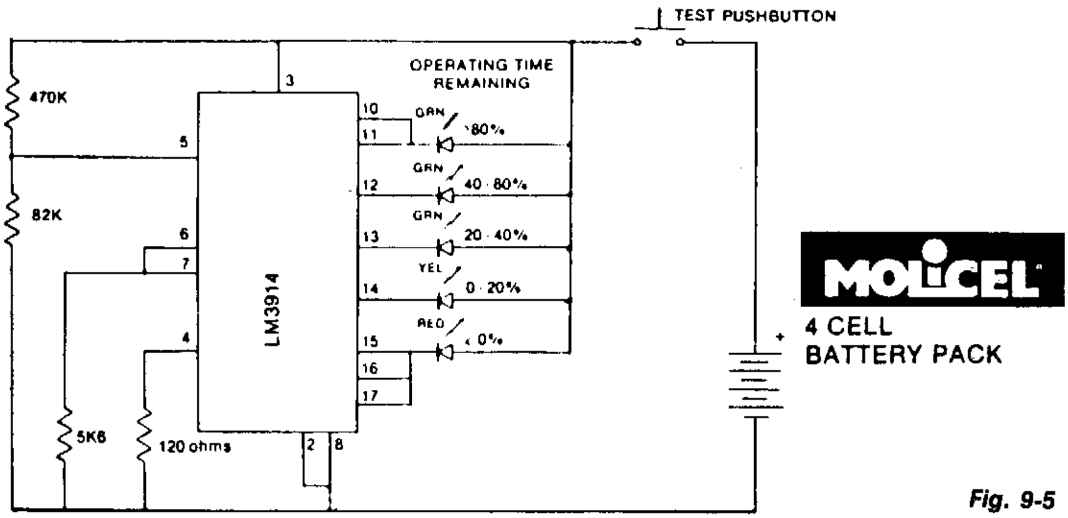


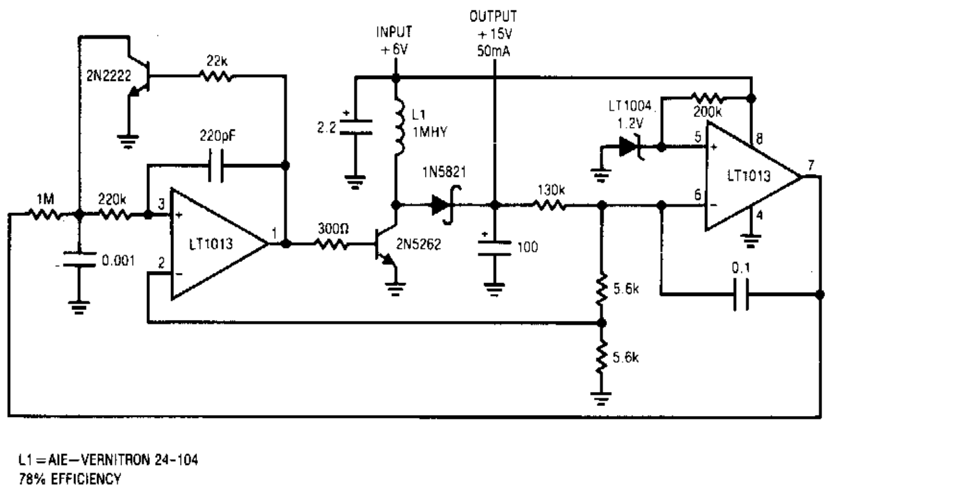
Fig. 9-5

MOLI ENERGY LIMITED

Circuit Notes

State-of-Charge indication of a sloping-voltage discharge can be used as a state-of-charge indicator. A typical voltage comparator circuit that gives a visual indication of state-of-charge is shown. Components identified are for a 4-cell input voltage of 9.6 to 5.2 volts.

STEP-UP SWITCHING REGULATOR FOR 6-V BATTERY

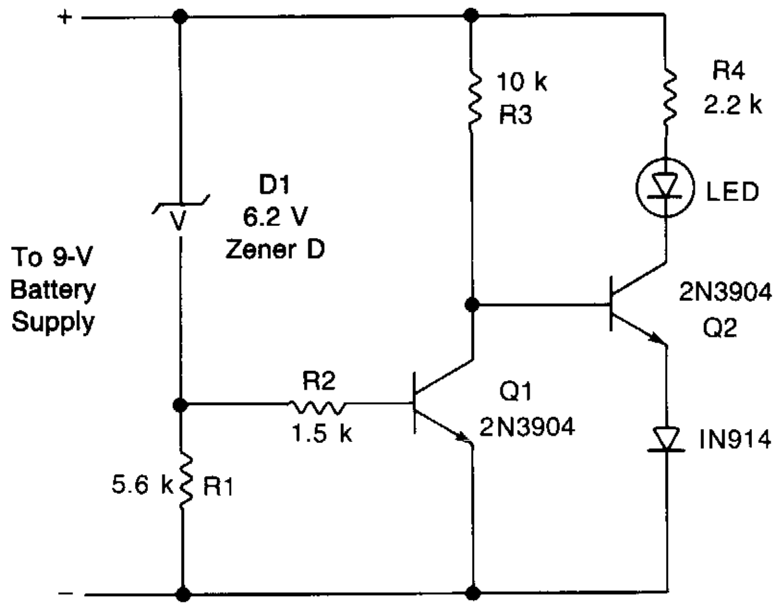


L1 = AIE—VERNITRON 24-104
78% EFFICIENCY

LINEAR TECHNOLOGY CORP.

Fig. 9-6

BATTERY VOLTAGE MONITOR



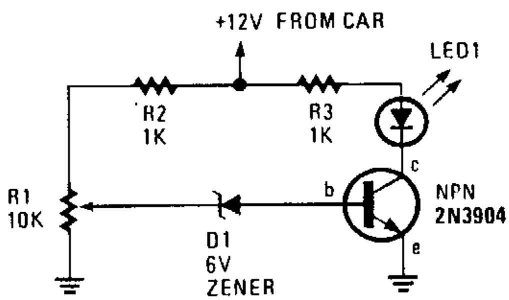
WILLIAM SHEETS

Fig. 9-7

Circuit Notes

This circuit gives an early warning of the discharge of batteries. Zener diode D1 is chosen for the voltage below which an indication is required (9 V). Should the supply drop to below 7 V, D1 will cease conducting causing Q1 to shut off. Its collector voltage will now increase causing Q2 to start conducting via LED1 and its limiting resistor R4.

BATTERY MONITOR



Circuit Notes

The circuit is quick and easy to put together and install, and tells you when battery voltage falls below the set limit as established by R1 (a 10,000-ohm potentiometer). It can indicate, via LED1, that the battery may be defective or in need of change if operating the starter causes the battery voltage to drop below the present limit.

TAB BOOKS, INC.

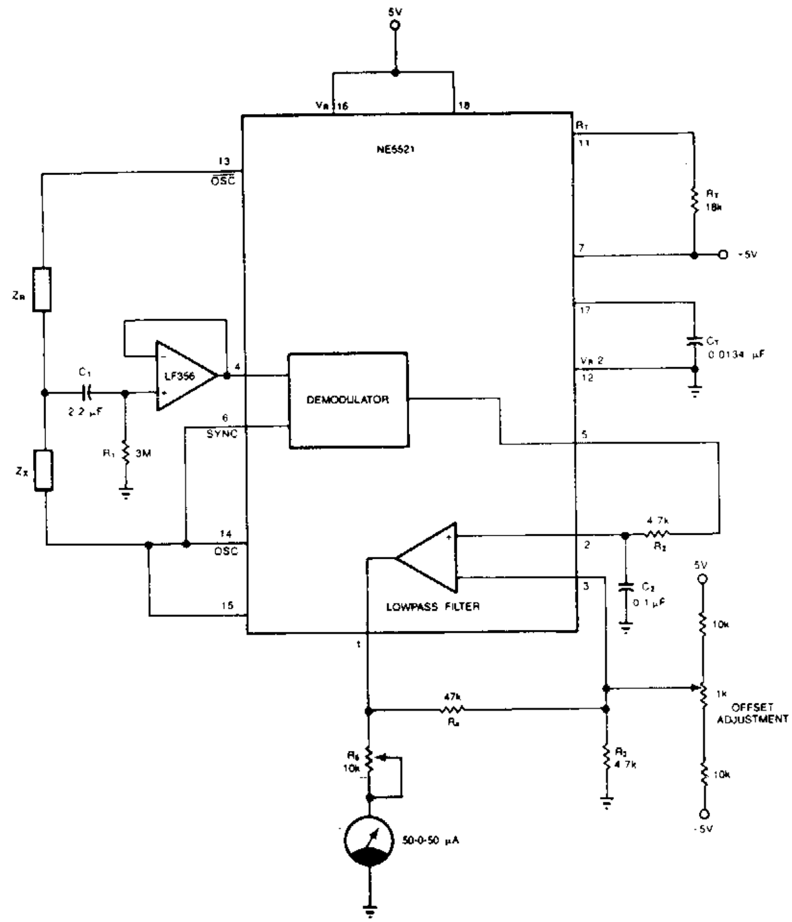
Fig. 9-8

10

Bridge Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AC BRIDGE



SIGNETICS *Fig. 10-1*

Circuit Notes

The circuit provides a simple and cost-effective solution to matching resistors and capacitors. Impedances Z_R and Z_X form a half-bridge, while OSC and $\overline{\text{OSC}}$ excite the bridge differentially. The external op amp is a FET input amplifier (LF356) with very low input bias current on the order of 30 pA (typical). C1 allows ac coupling by blocking the dc common mode voltage from the bridge, while R1 biases the output of LF356 to 0 V at dc. Use of FET input op amp insures that dc offset due to bias current through R1 is negligible. Ac output of the demodulator is filtered via the uncommitted amp to provide dc voltage for the meter. The 10 k potentiometer, R5, limits the current into the meter to a safe level. Calibration begins by placing equal impedances at Z_R and Z_X, and the system offset is nulled by the offset adjust circuit so that Pin 1 is at 0 V. Next, known values are placed at Z_X and the meter deviations are calibrated. The bridge is now ready to measure an unknown impedance at Z_X with $\pm 0.05\%$ accuracy or better.

BRIDGE-BALANCE INDICATOR

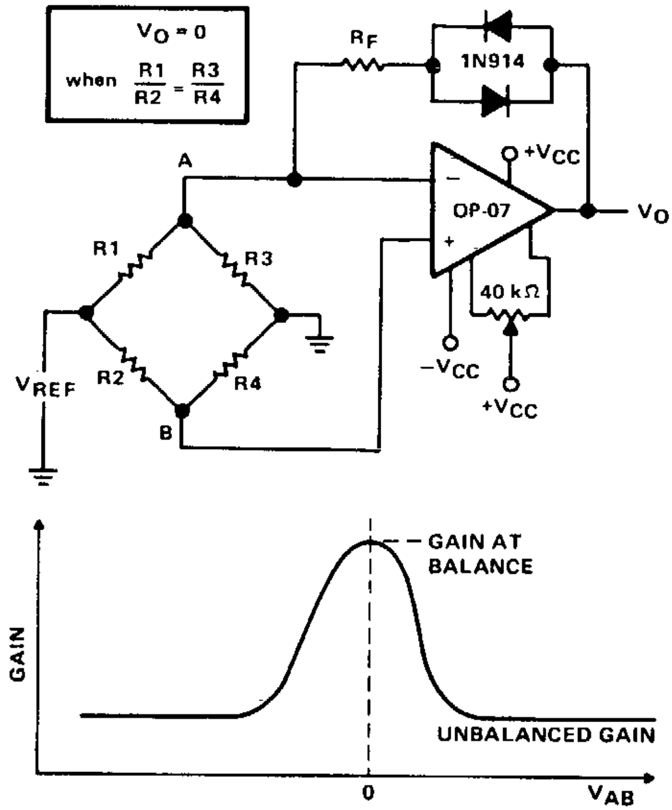


Fig. 10-2

TEXAS INSTRUMENTS

Circuit Notes

Indicator provides an accurate comparison of two voltages by indicating their degree of balance (or imbalance). Detecting small variations near the null point is difficult with the basic Wheatstone bridge alone. Amplification of voltage differences near the null point will improve circuit accuracy and ease of use.

The 1N914 diodes in the feedback loop result in high sensitivity near the point of balance (R1/R2 = R3/R4). When the bridge is unbalanced the amplifier's closed-loop gain is approximately R_F/r, where r is the parallel equivalent of R1 and R3. The resulting gain equation is $G = R_F(1/R1 + 1/R3)$. During an unbalanced condition the voltage at point A is different from that at point B. This difference voltage (V_{AB}), amplified by the gain factor G, appears as an output voltage, As the bridge approaches a balanced condition (R1/R2 = R3/R4), V_{AB} approaches zero. As V_{AB} approaches zero the 1N914 diodes in the feedback loop lose their forward bias and their resistance increases, causing the total feedback resistance to increase. This increases circuit gain and accuracy in detecting a balanced condition. The figure shows the effect of approaching balance on circuit gain. The visual indicator used at the output of the OP-07 could be a sensitive voltmeter or oscilloscope.

BRIDGE CIRCUIT

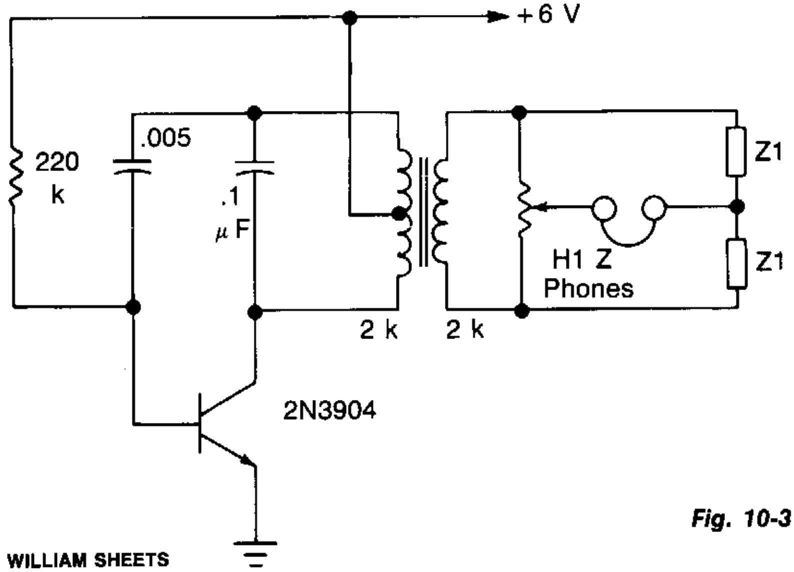


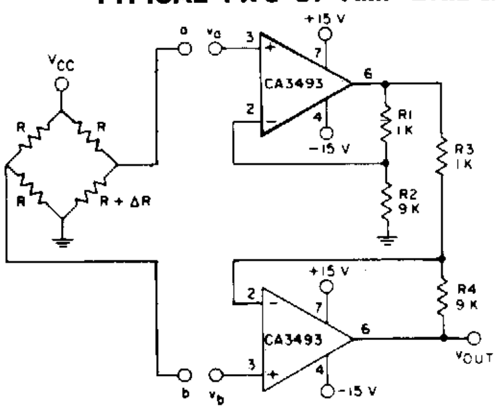
Fig. 10-3

WILLIAM SHEETS

Circuit Notes

The transistor is connected as an audio oscillator, using an audio transformer in the collector. The secondary goes to a linear pot. The ratio between the two parts of the pot from the slider is proportional to the values of Z1 and Z2 when no signal is heard in the phones.

TYPICAL TWO OP AMP BRIDGE-TYPE DIFFERENTIAL AMPLIFIER



$V_{OUT} = V_b - V_a \left(\frac{R4}{R3} + 1\right)$

$A = \frac{V_{OUT}}{V_b - V_a} = \left(\frac{R4}{R3} + 1\right)$

FOR VALUES ABOVE $V_{OUT} = (V_b - V_a)(10)$

ALL RESISTANCE VALUES ARE IN OHMS

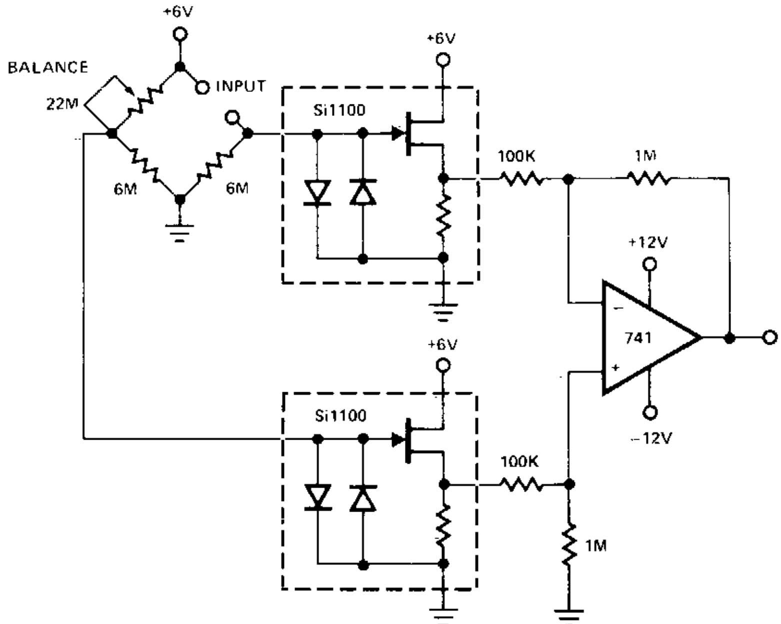
GENERAL ELECTRIC/RCA

Circuit Notes

Using a CA3493 BiMOS op amp to provide high input impedance and good common-mode rejection ratio (depends primarily on matching of resistor networks).

Fig. 10-4

LOW-POWER COMMON SOURCE AMPLIFIER



SILICONIX, INC.

Fig. 10-5

Circuit Notes

A circuit that will operate in the 10- to 20- microamp range at a 12-volt supply voltage. The diode protection is available in this configuration. The circuit voltage gain will be between 10 and 20, with extremely low power consumption (approximately 250 μW). This is very desirable for remote or battery operation where minimum maintenance is important.

AMPLIFIER FOR BRIDGE TRANSDUCERS

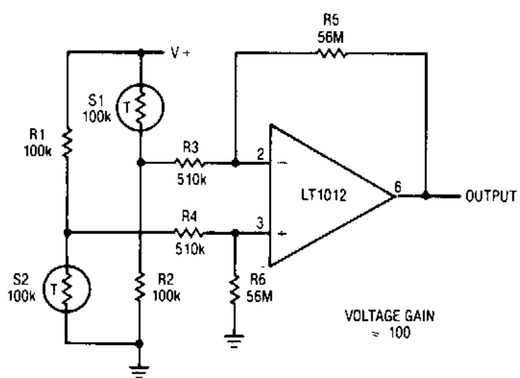
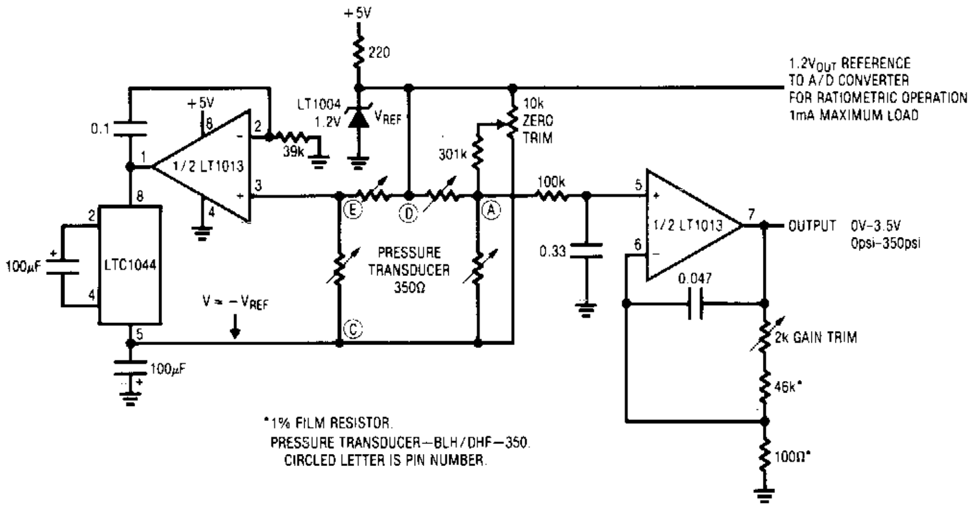


Fig. 10-6

LINEAR TECHNOLOGY CORP.

STRAIN GAUGE BRIDGE SIGNAL CONDITIONER



LINEAR TECHNOLOGY CORP.

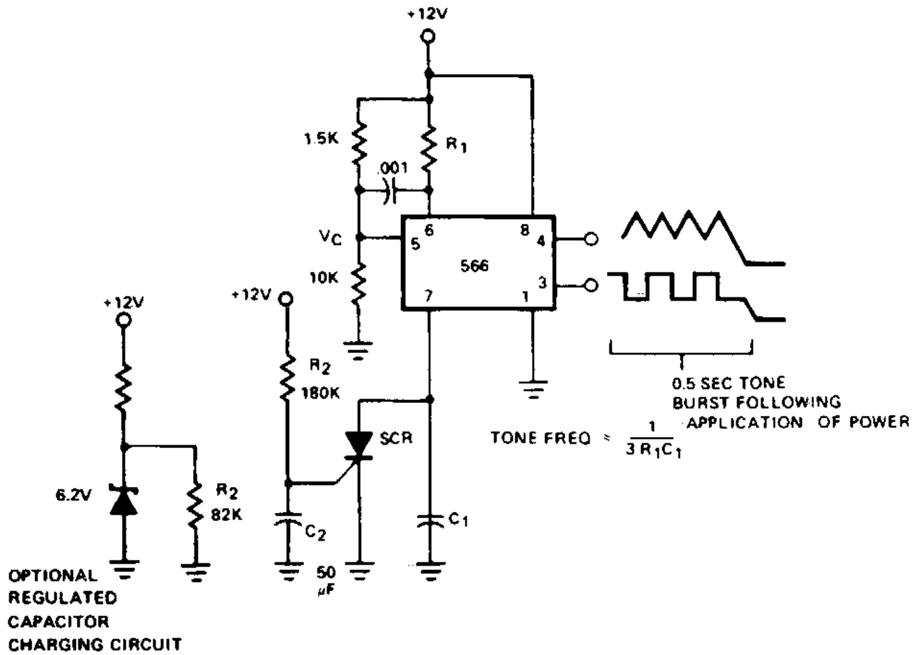
Fig. 10-7

11

Burst Generators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SINGLE-TONE BURST GENERATOR



SIGNETICS

Fig. 11-1

Circuit Notes

The tone burst generator supplies a tone for one-half second after the power supply is activated; its intended use is a communications network alert signal. Cessation of the tone is accomplished at the SCR, which shunts the timing capacitor C1 charge current when activated. The SCR is gated on when C2 charges up to the gate voltage which occurs in 0.5 seconds. Since only 70 μA are available for triggering, the SCR must be sensitive enough to trigger at this level. The triggering current can be increased, of course, by reducing R2 (and increasing C2 to keep the same time constant). If the tone duration must be constant under widely varying supply voltage conditions, the optional Zener diode regulator circuit can be added, along with the new value for R_2 $R_2{}' = 82$ kΩ. If the SCR is replaced by an npn transistor, the tone can be switched on and off at will at the transistor base terminal.

SQUARE WAVEFORM MULTIBURST GENERATOR

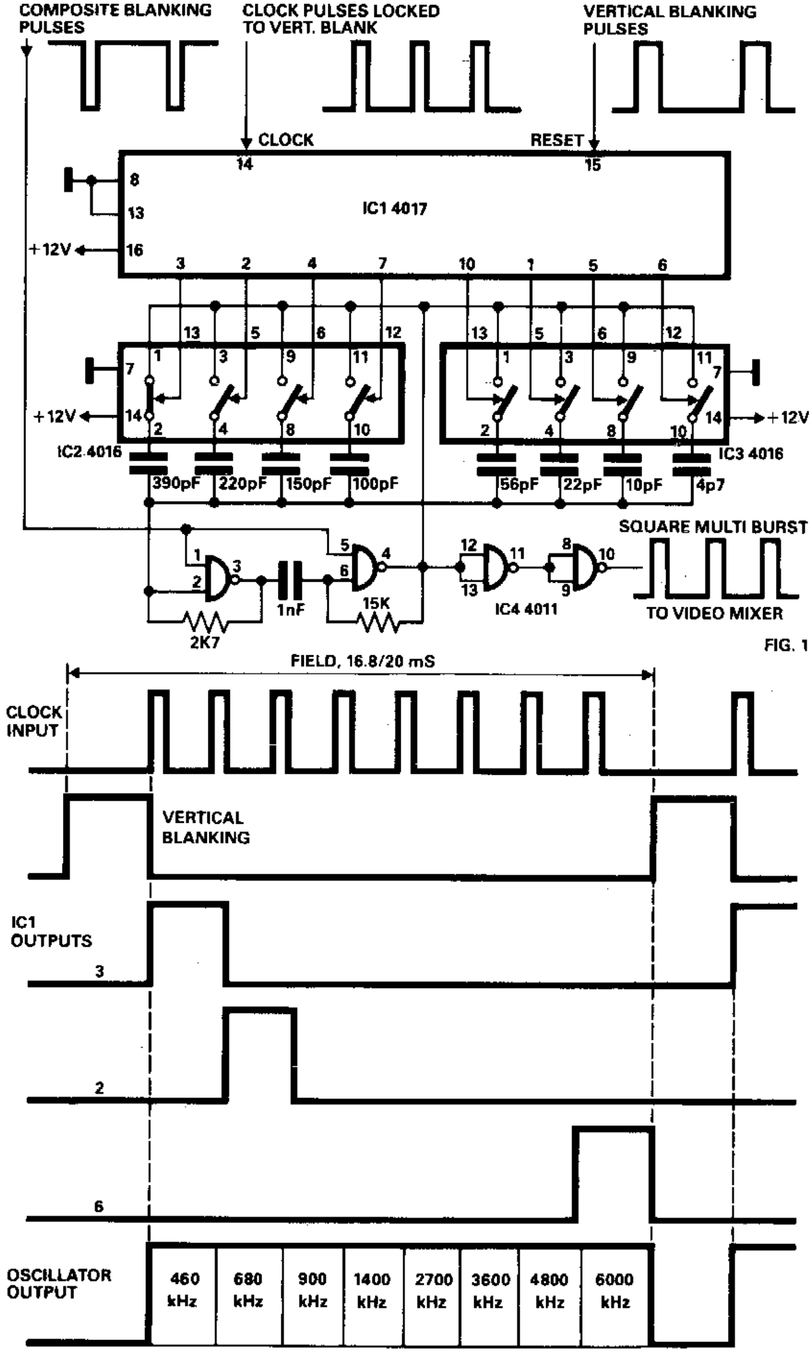


FIG. 1

ELECTRONIC ENGINEERING

Fig. 11-2

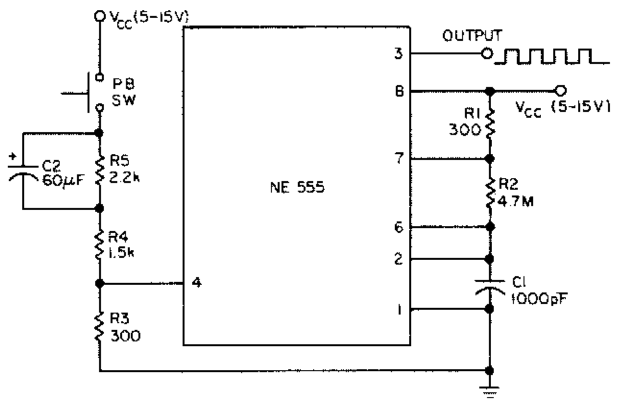
SQUARE WAVEFORM MULTIBURST GENERATOR, Continued.

Circuit Notes

The generator described here is intended for multiburst signal square waveform generation and can be used as a device for characterizing the response of TV monitor amplifiers as shown. The circuit is an RC oscillator with NAND gates (IC4-4011), with its capacitor C changed periodically by means of bilateral switches (IC2, IC3-4016). The control inputs of bilateral switches are driven by the outputs of a counter/decoder (IC1-4017) the operation of which is determined by generated clock pulses, so that they occur eight times at half-picture (field). These pulses are locked to vertical blank pulses.

Horizontal synchronization is achieved by means of composite blanking pulses (negative polarization) applied to pins 1 and 5 of IC4. The oscillator frequency changes in the following discrete steps: 460 kHz, 680 kHz, 900 kHz, 1400 kHz, 2700 kHz, 3600 kHz, for the time of one frame. The video signal is fed on a mixer where it is superimposed with a composite sync signal.

SINGLE-TIMER IC PROVIDES SQUARE-WAVE TONE BURSTS



ELECTRONIC DESIGN

Fig. 11-3

Circuit Notes

The tone-burst generator gives a 50-ms burst of 1.5 kHz square waves with each operation of the pushbutton and can source or sink 200 mA.

STROBE-TONE BURST GENERATOR

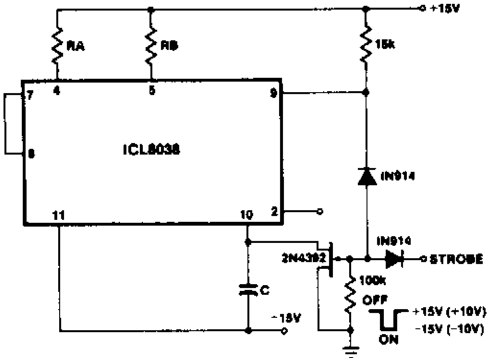
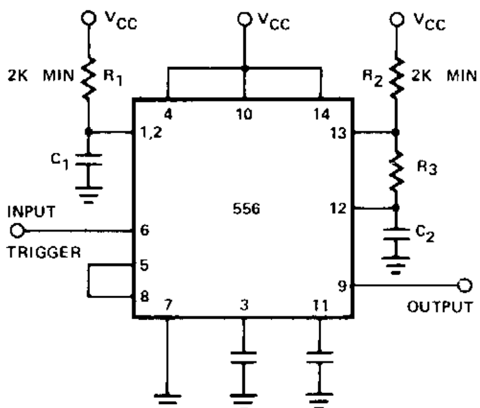


Fig. 11-4

INTERSIL

TONE BURST GENERATOR



Circuit Notes

The dual timer makes an excellent tone burst generator. The first half is connected as a one shot and the second half as an oscillator. The pulse established by the one shot turns on the oscillator allowing a burst of pulses to be generated.

SIGNETICS

Fig. 11-5

12

Capacitance Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

CAPACITANCE-TO-VOLTAGE METER

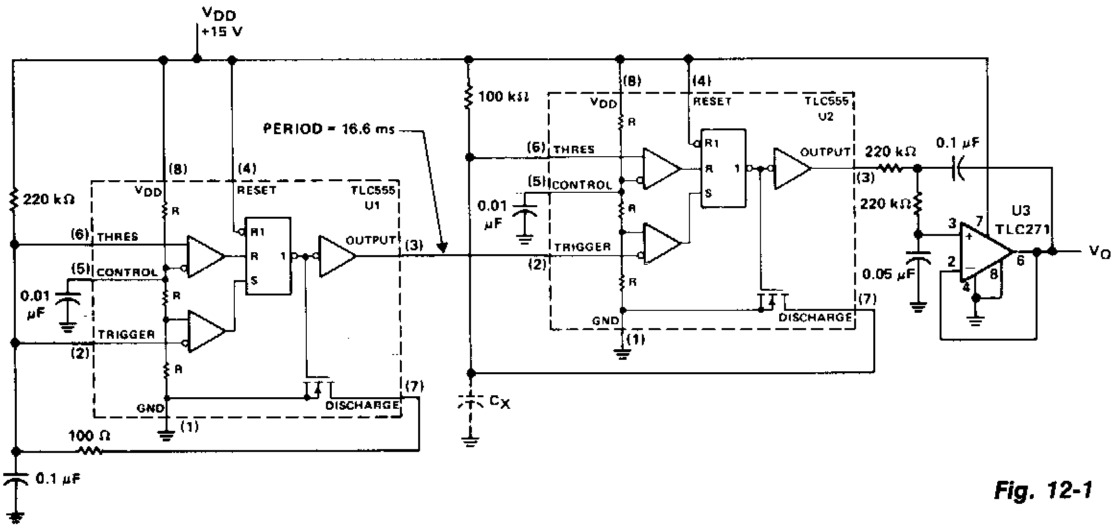


Fig. 12-1

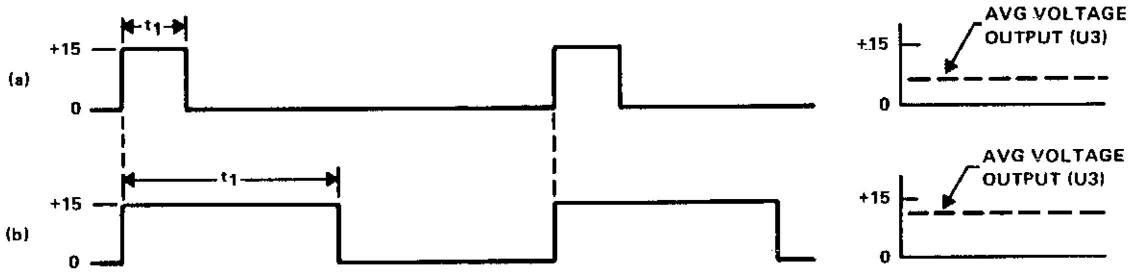
TEXAS INSTRUMENTS

Circuit Notes

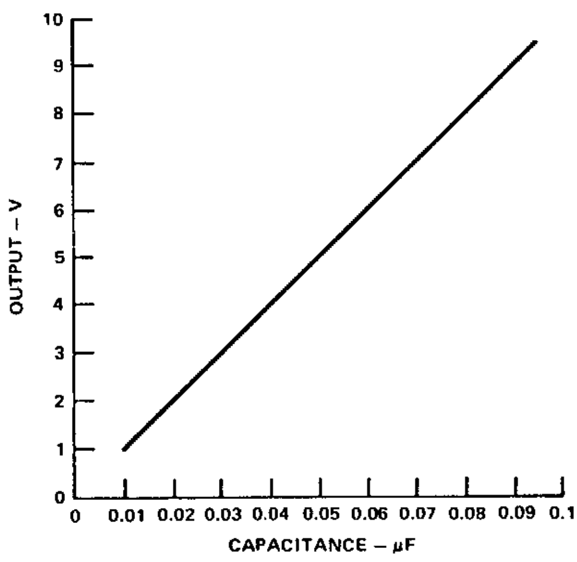
Timer U1 operates as a free-running oscillator at 60 Hz, providing trigger pulses to timer U2 which operates in the monostable mode. Resistor R1 is fixed and capacitor Cx is the capacitor being measured. While the output of U2 is 60 Hz, the duty cycle depends on the value of Cx. U3 is a combination low-pass filter and unity-gain follower whose dc voltage output is the time-averaged amplitude of the output pulses of U2, as shown in the timing diagram.

The diagram shows when the value of Cx is small the duty cycle is relatively low. The output pulses are narrow and produce a lower average dc voltage level at the output of U3. As the capacitance value of Cx increases, the duty cycle increases making the output pulses at U2 wider and the average dc level output at U3 increases. The graph illustrates capacitance values of 0.01 µF to 0.1 µF plotted against the output voltage of U3. Notice the excellent linearity and direct one-to-one scale calibration of the meter. If this does not occur the 100 k ohm resistor, R1, can be replaced with a potentiometer which can be adjusted to the proper value for the meter being used.

CAPACITANCE-TO-VOLTAGE METER (CONT'D)



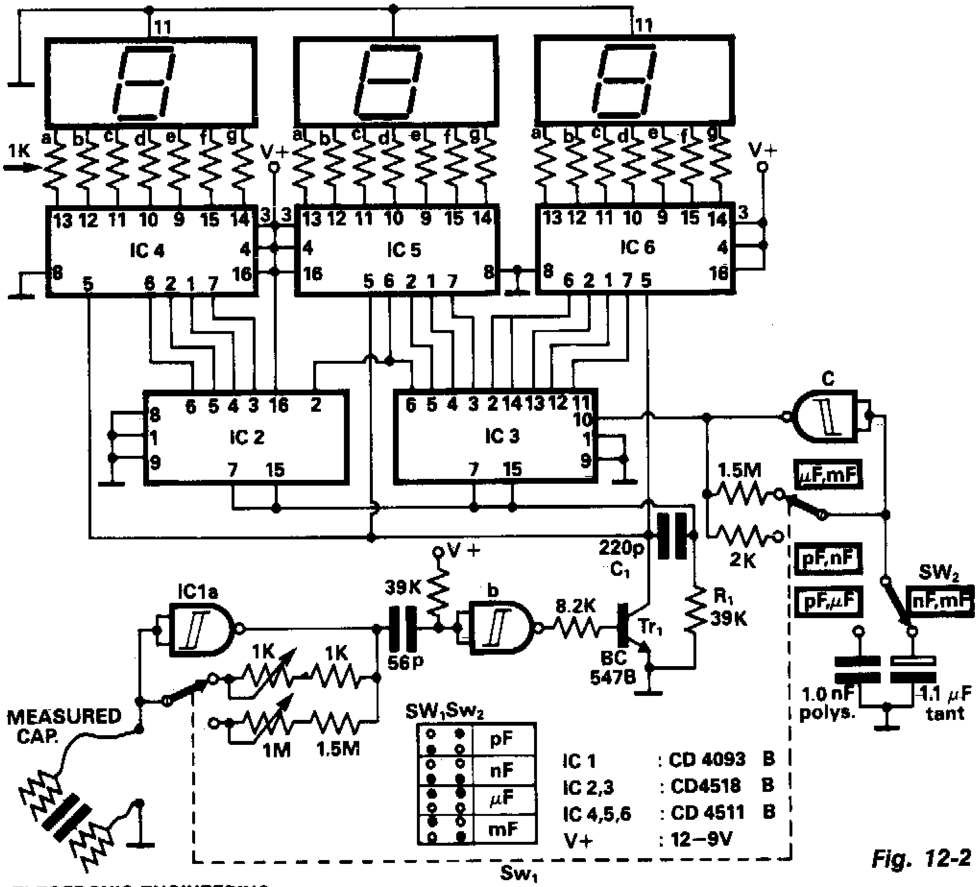
U2 Duty Cycle Change



TEXAS INSTRUMENTS

Fig. 12-1

ACCURATE DIGITAL CAPACITANCE METER



ELECTRONIC ENGINEERING

Fig. 12-2

Circuit Notes

The principle of operation is counting the pulse number derived from a constant frequency oscillator during a fixed time interval produced by another lower frequency oscillator. This oscillator uses the capacitor being measured as the timing. The capacitance measurement is proportional during pulse counting during a fixed time interval. The astable oscillator formed by IC1c produces a pulse train of constant frequency. Gate IC1a also forms an oscillator whose oscillation period is given approximately by the equation: $T = 0.7\ RC$.

Period T is linearly dependent on the capacitance C. This period is used as the time interval for one measurement. The differentiator network following the oscillator creates the negative spikes shaped in narrow pulses by IC1b NAND Schmitt Trigger. The differentiator formed by R1 and C1 produces a negative spike which resets the counters. The display shows the number of high frequency oscillator pulses entering the counter during the measurement period.

13

Circuit Protection Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

OVERVOLTAGE PROTECTOR

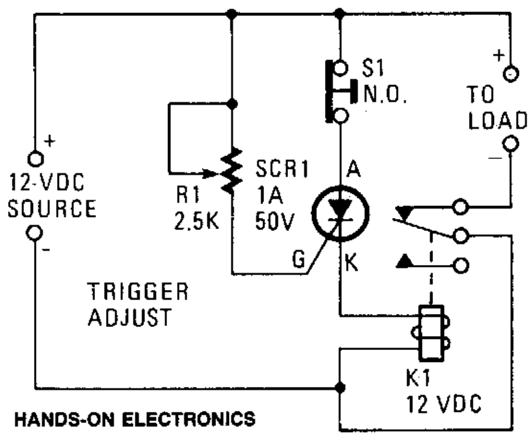


Fig. 13-1

Circuit Notes

A silicon-controlled rectifier is installed in parallel with the 12-V line and connected to a normally-closed 12-V relay, K1. The SCR's gate circuit is used to sample the applied voltage. As long as the applied voltage stays below a given value, SCR1 remains off and K1's contacts remain closed, thereby supplying power to the load. When the source voltage rises above 12 V, sufficient current is applied to the gate of SCR1 to trigger it into conduction. The trigger point of SCR1 is dependent on the setting of R1. Once SCR1 is triggered (activating the relay), K1's contacts open, halting current flow to the load.

HIGH SPEED ELECTRONIC CIRCUIT BREAKER

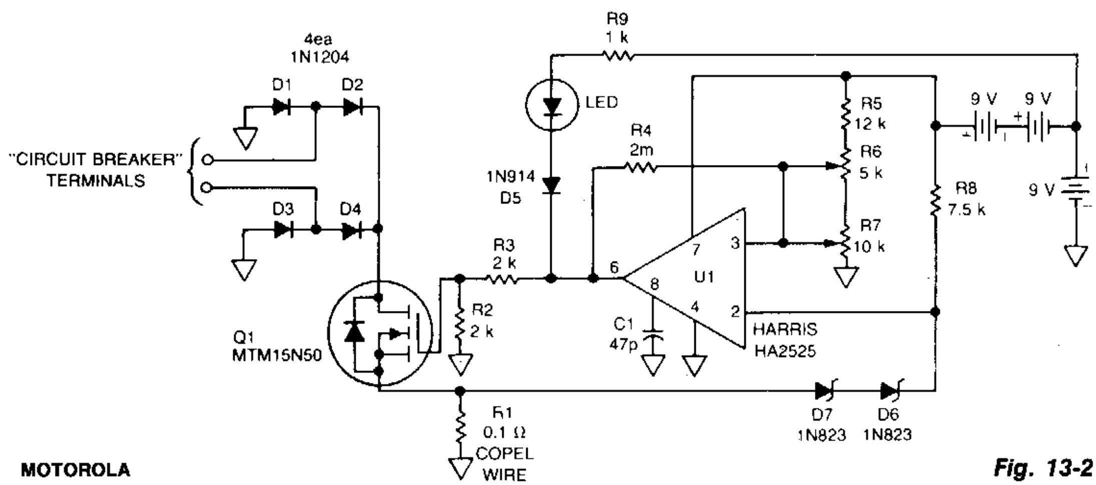


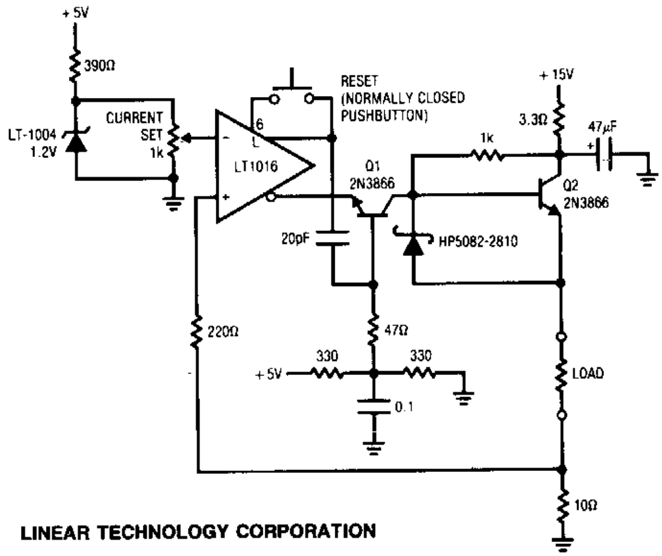
Fig. 13-2

Circuit Notes

This 115 Vac, electronic circuit breaker uses the low drive power, low on resistance and fast turn off of the TMOS MTM15N50. The trip point is adjustable, LED fault indication is provided and battery power provides complete circuit isolation.

The two "circuit breaker" terminals are across one leg of a full wave diode bridge consisting of D1-D4. Normally, Q1 is turned ON so that the circuit breaker looks like a very low resistance. One input to comparator U1 is a fraction of the internal battery voltage and the other input is the drop across zeners D6 and D7 and the voltage drop across R1. If excessive current is drawn, the voltage drop across R1 increases beyond the comparator threshold (determined by the setting of R6), U1 output goes low, Q1 turns OFF, and the circuit breaker "opens." When this occurs, the LED fault indicator is illuminated.

12 ns CIRCUIT BREAKER



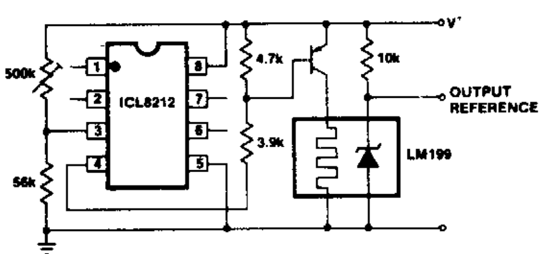
LINEAR TECHNOLOGY CORPORATION

Fig. 13-3

Circuit Notes

This circuit will turn off current in a load 12 ns after it exceeds a preset value. Under normal conditions the voltage across the 10 ohm shunt is smaller than the potential at the LT1016's negative input. This keeps Q1 off and Q2 receives bias, driving the load. When an overload occurs the current through the 10 ohm sense resistor begins to increase. When this current exceeds the preset value, the LT1016's outputs reverse. This provides ideal turn-on drive for Q1 and it cuts off Q2 in 5 ns. The delay from the onset of excessive load current to complete shutdown is just 13 ns. Once the circuit has triggered, the LT1016 is held in its latched state by feedback from the non-inverting output. When the load fault has been cleared the pushbutton can be used to reset the circuit.

LOW VOLTAGE POWER DISCONNECTOR



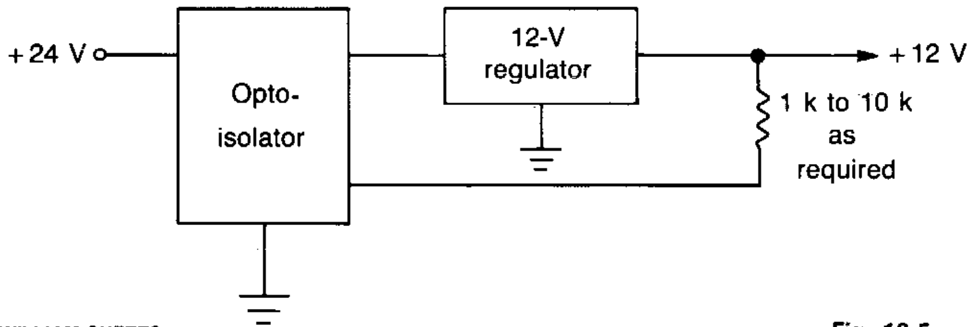
INTERSIL

Fig. 13-4

Circuit Notes

There are some classes of circuits that require the power supply to be disconnected if the power supply voltage falls below a certain value. As an example, the National LM199 precision reference has an on chip heater which malfunctions with supply voltages below 9 volts causing an excessive device temperature. The ICL8212 can be used to detect a power supply voltage of 9 volts and turn the power supply off to the LM199 heater section below that voltage.

AUTOMATIC POWER-DOWN PROTECTION CIRCUIT



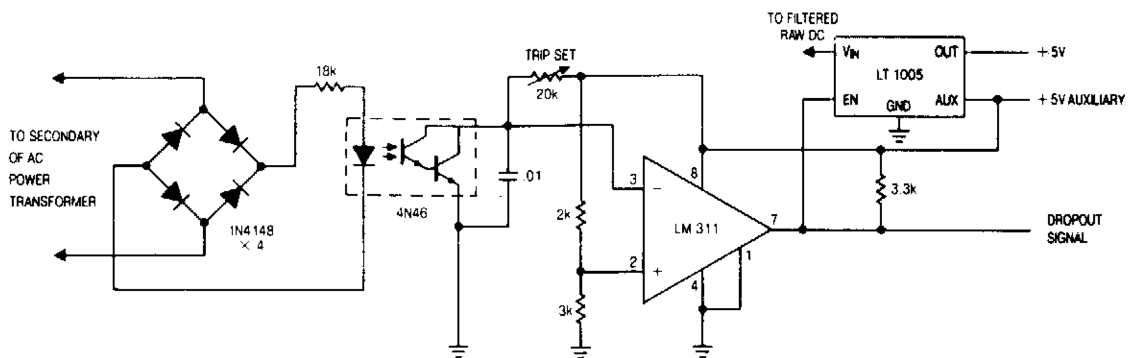
WILLIAM SHEETS

Fig. 13-5.

Circuit Notes

This circuit is faster than a fuse and automatically resets itself when a short is removed. The normal regulated dc input line is opened and the phototransistor of the opto isolator is connected in series with the source and regulator. Between the output of the regulator and ground is a LED and an associated current-limiting resistor, placed physically close to the surface of the photosensitive device. As long as the regulator is delivering its rated output, the LED glows and causes the photo device to have a low resistance. Full current is thus allowed to flow. If a short circuit occurs on the output side of the regulator, the LED goes dark, the resistance of the photo device increases, and the regulator shuts off. When the short is removed, the LED glows, and the regulator resumes operation.

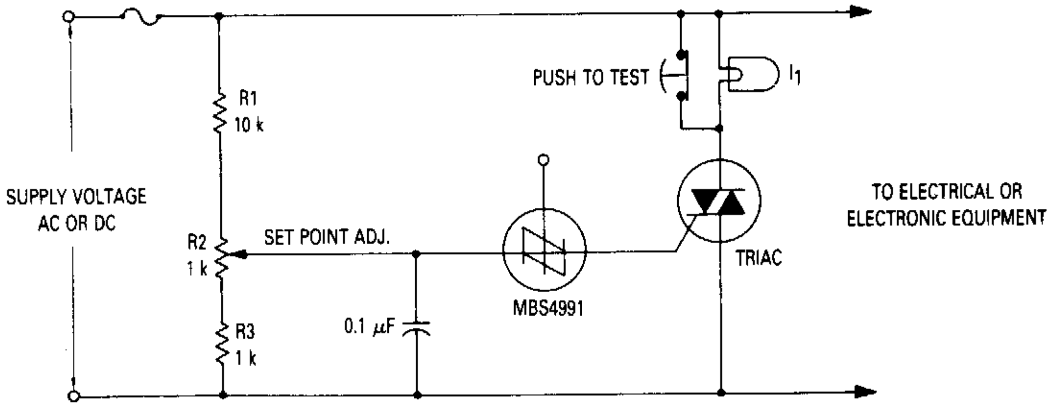
LINE DROPOUT DETECTOR



LINEAR TECHNOLOGY CORP.

Fig. 13-6

ELECTRONIC CROWBAR



MOTOROLA *Fig. 13-7*

Circuit Notes

Where it is desirable to shut down equipment rather than allow it to operate on excessive supply voltage, an electronic ''crowbar'' circuit can be employed to quickly place a short-circuit across the power lines, thereby dropping the voltage across the protected device to near zero and blowing a fuse. Since the TRIAC and SBS are both bilateral devices, the circuit is equally useful on ac or dc supply lines. With the values shown for R1, R2 and R3, the crowbar operating point can be adjusted over the range of 60 to 120 volts dc or 42 to 84 volts ac. The resistor values can be changed to cover a different range of supply voltages. The voltage rating of the TRIAC must be greater than the highest operating point as set by R2. I_1 is a low power incandescent lamp with a voltage rating equal to the supply voltage. It may be used to check the set point and operation of the unit by opening the test switch and adjusting the input or set point to fire the SBS.

14

Clock Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

THREE PHASE CLOCK FROM A REFERENCE CLOCK

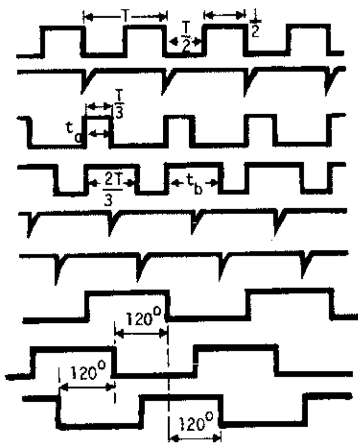
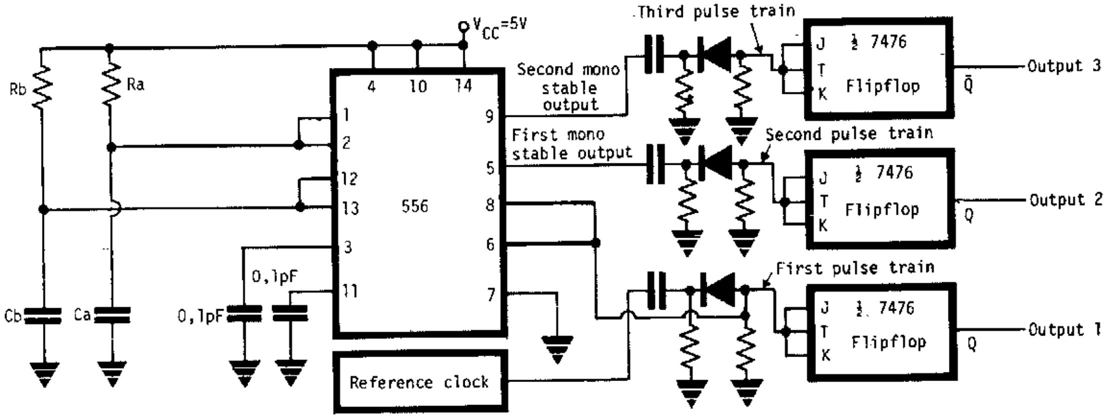


Fig. 14-1

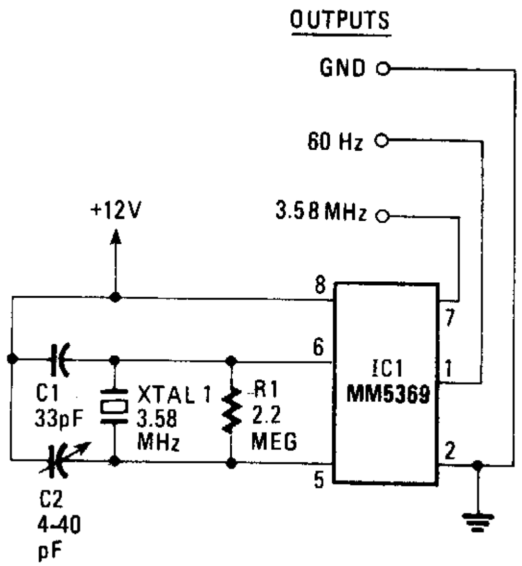


ELECTRONIC ENGINEERING

Circuit Notes

The circuit provides three square wave outputs with 120° of phase difference between each other. Reference clock frequency is twice that of the required frequency. This can be obtained from a crystal oscillator with a chain of dividers or by using LM 555 in 50% duty cycle astable mode. If 1/T is the frequency of the reference clock, the dual timer 556 is connected to give two mono-stable output pulses of duration T/3 and 2T/3. The first timer R and C value are adjusted so that $t_a = 1.1 R_a C_a = T/3$ and the second timer R and C values so that $t_b = 1.1 R_b C_b = 2T/3$. For triggering the two monostables a negative pulse train (1st) is derived from the reference clock with a differentiator and a clipper combination as shown. The three pulse trains trigger three JK flip flops giving three phase square wave outputs.

60 Hz CLOCK PULSE GENERATOR



HANDS-ON ELECTRONICS

Fig. 14-2

Circuit Notes

The circuit provides a clean, stable square wave and it will operate on anywhere from 6 to 15 volts. The IC and color-burst crystal are the kind used in TV receivers. The 3.58 MHz output makes a handy marker signal for shortwave bands.

15

Comparators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW-POWER COMPARATOR WITH LESS THAN 10 μV HYSTERESIS

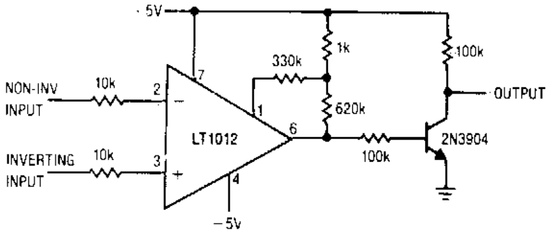
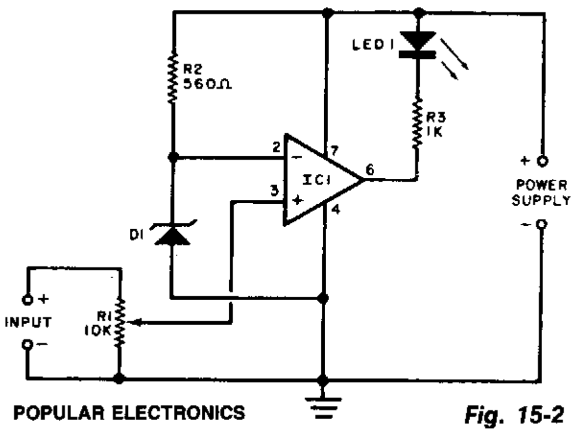


Fig. 15-1

LINEAR TECHNOLOGY CORP.

VOLTAGE MONITOR/COMPARATOR



POPULAR ELECTRONICS

Fig. 15-2

Circuit Notes

A portion of the monitored voltage (determined by R1's adjustment) is compared to a fixed voltage obtained from a zener reference network, R2-D1. As long as the monitored voltage remains at or above its present monitor point (determined by R1's setting), the output indicator, LED1, remains dark. If the voltage drops below this level, the LED goes on. D1 is a 3.3-V zener. A 12 Vdc power supply is suitable for monitoring input voltages of up to 12 volts.

LIMIT COMPARATOR

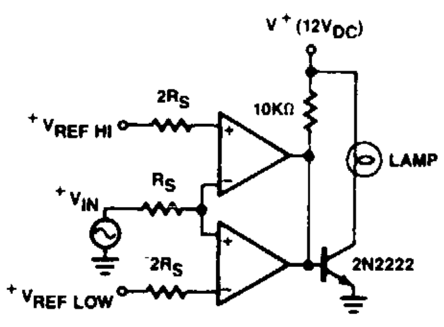
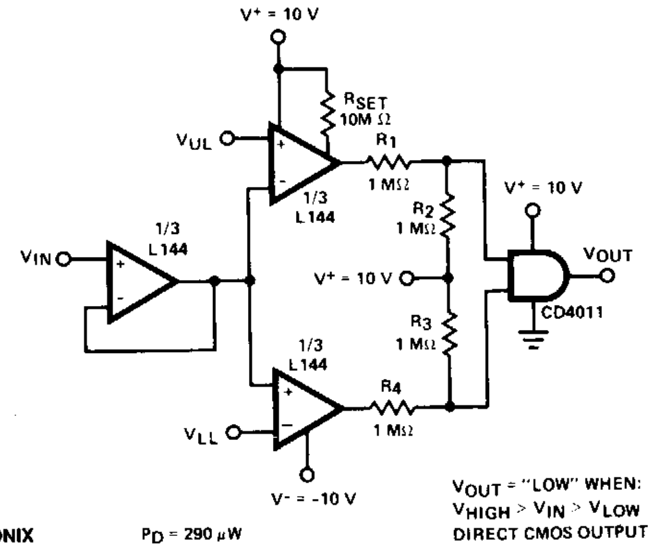


Fig. 15-3

SIGNETICS

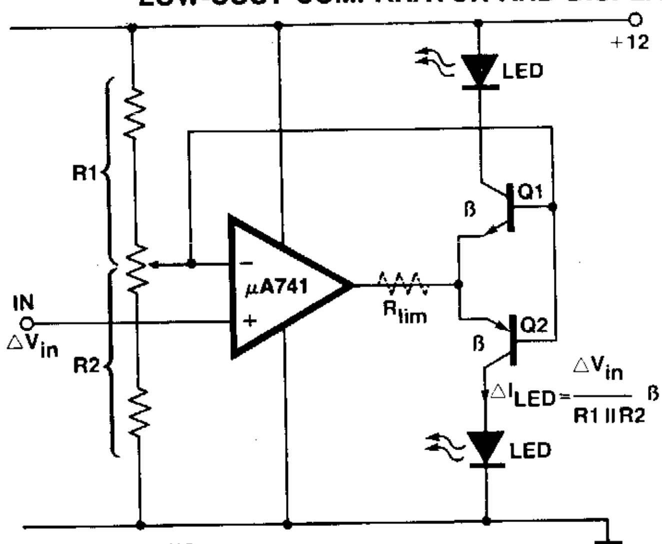
DOUBLE-ENDED LIMIT COMPARATOR



SILICONIX

Fig. 15-4

LOW-COST COMPARATOR AND DISPLAY



ELECTRONIC ENGINEERING

Fig. 15-5

Circuit Notes

An op amp is used as a comparator and a sink for LED current. The output voltage of the amplifier changes about 1.4 V depending on the direction of the current. Only one transistor is on at any time. Maximum LED current is limited to 25 mA by overcurrent protection of the uA741. If LEDs are not capable of carrying such a current or an alternative op amp is used and an additional resistor R_{lim} is necessary.

WINDOW COMPARATOR

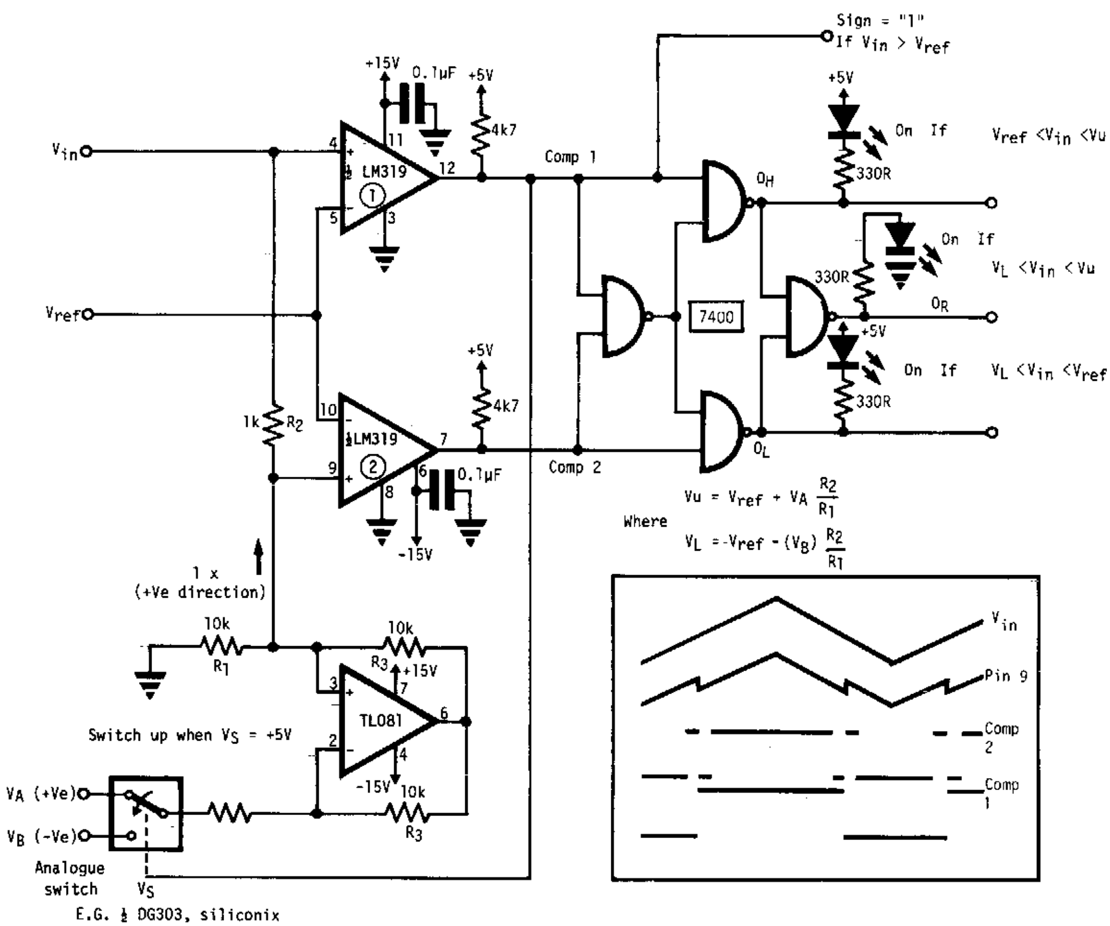


Fig. 15-6

ELECTRONIC ENGINEERING

Circuit Notes

This circuit provides independently adjustable upper and lower threshold settings, and has sign, in window range, in upper window, and in lower window digital outputs.

COMPARATOR DETECTS POWER SUPPLY OVERVOLTAGES, CATCHES GLITCHES

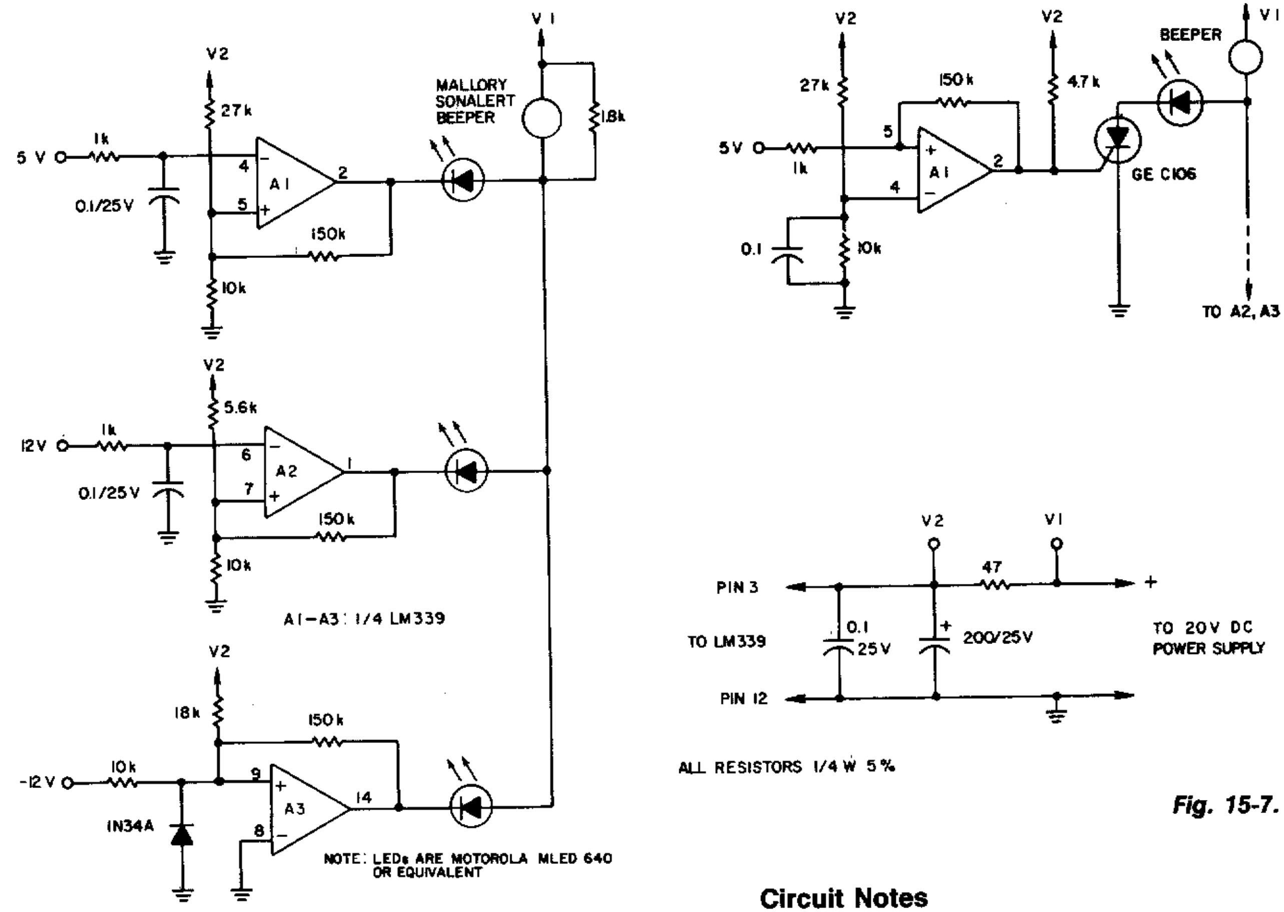


Fig. 15-7.

ELECTRONIC DESIGN

Circuit Notes

(A) To maintain an alarm condition when an overvoltage transient disappears, add an SCR to the comparator circuits. For SCR operation, voltages to the comparator inputs are inverted. (B) The triple-voltage monitoring circuit detects transient power-supply overvoltages. If excessive voltage momentarily appears at the 5, 12 and −12 V inputs, the LED for that circuit lights and the beeper sounds for as long as the overvoltage lasts.

HIGH-LOW LEVEL COMPARATOR WITH ONE OP AMP

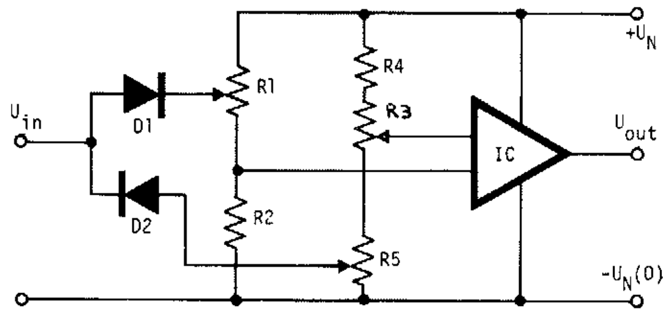


Fig. 15-8

ELECTRONIC ENGINEERING

Circuit Notes

The voltage to be compared is fed through diode D1 and D2 to the voltage dividers R1 and R5 where the low and high limits are present. When the voltage level of an input signal exceeds the high threshold limit set with potentiometer R1, the diode D1 becomes forward biased and the increased voltage on the inputs of the op amp drives it into positive saturation. Similarly, a decrease of the input voltage at the op amp inputs turns the op amp to positive saturation. Potentiometer R3 is used for zeroing the op amp in the off state.

HIGH-INPUT-IMPEDANCE WINDOW COMPARATOR

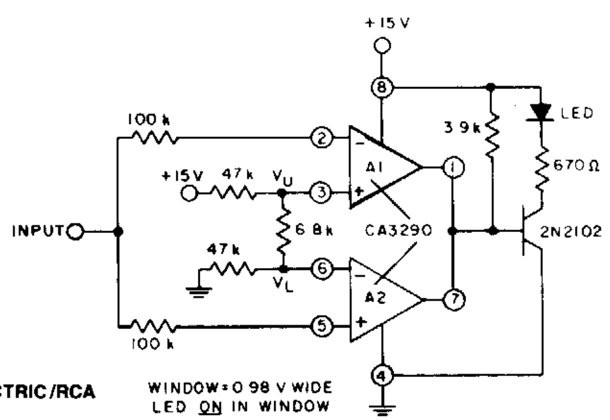


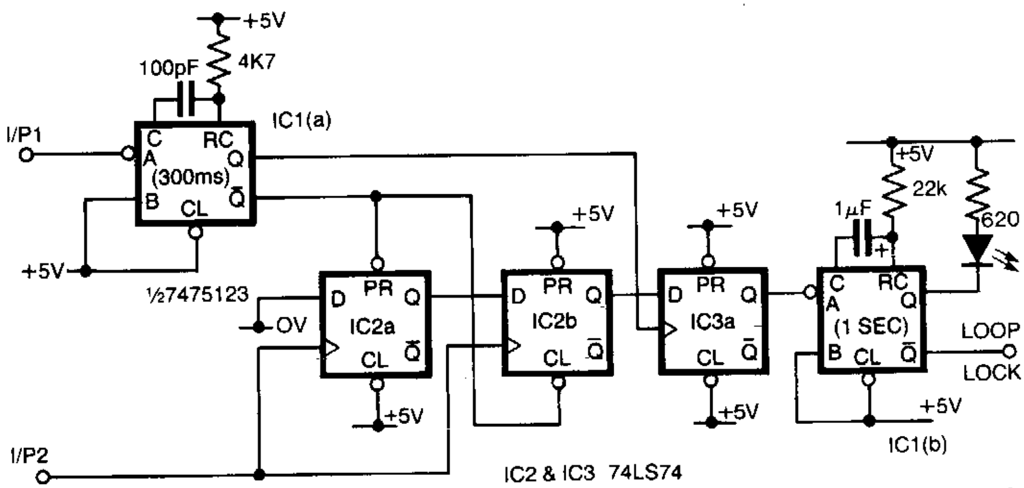
Fig. 15-9

GENERAL ELECTRIC/RCA

Circuit Notes

The circuit uses both halves of the CA3290 BiMOS dual voltage comparator. The LED will be turned "ON" whenever the input signal is above the lower limit (V_L) but below the upper limit (V_U).

FREQUENCY COMPARATOR



ELECTRONIC ENGINEERING

Fig. 15-10

Circuit Notes

Input 1 is used as a gating period, during which a single rising edge on input 2 will cause a logic 1 output-any other number, indicating non-identical frequencies causes a logic 0 output.

IC1a converts input 1 to a narrow pulse which initializes IC2 which forms a two-stage shift register clocked by input 2. On the first edge of input 2 a logic 1 appears on the output of IC2b and for all subsequent inputs a logic 0 is present. At the end of the gating period this output is latched by IC3 forming the lock output. As this is only valid for one input period a monostable is added to the output to enable, for example, visual monitoring of the output. Either output from IC3 can be used depending on which state is most important. As connected the failure state is indicated.

DEMONSTRATION COMPARATOR CIRCUIT

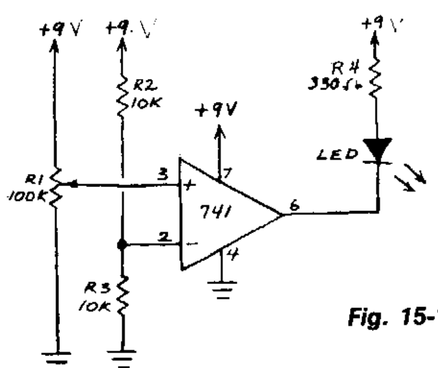


Fig. 15-11

POPULAR ELECTRONICS

Circuit Notes

This circuit is an op amp without a feedback resistor. R2 and R3 junction point sets the reference voltage. When the input voltage set by R1 is below the reference voltage the LED glows. If voltage is above reference, the LED goes off.

LED FREQUENCY COMPARATOR

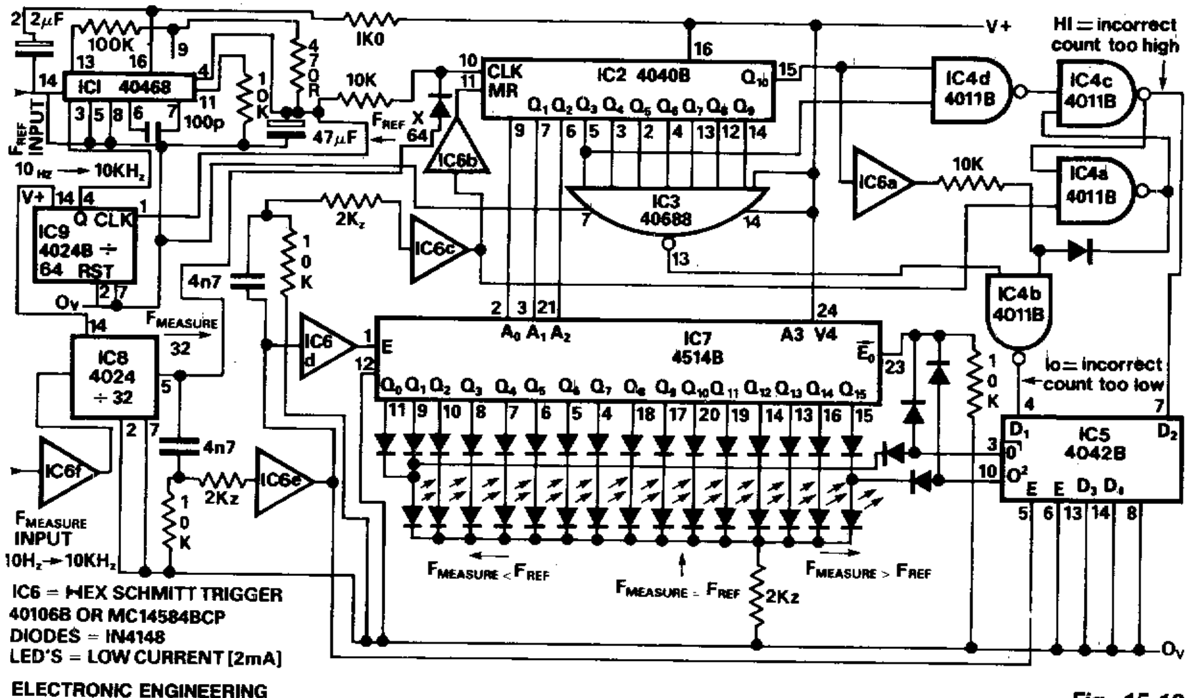


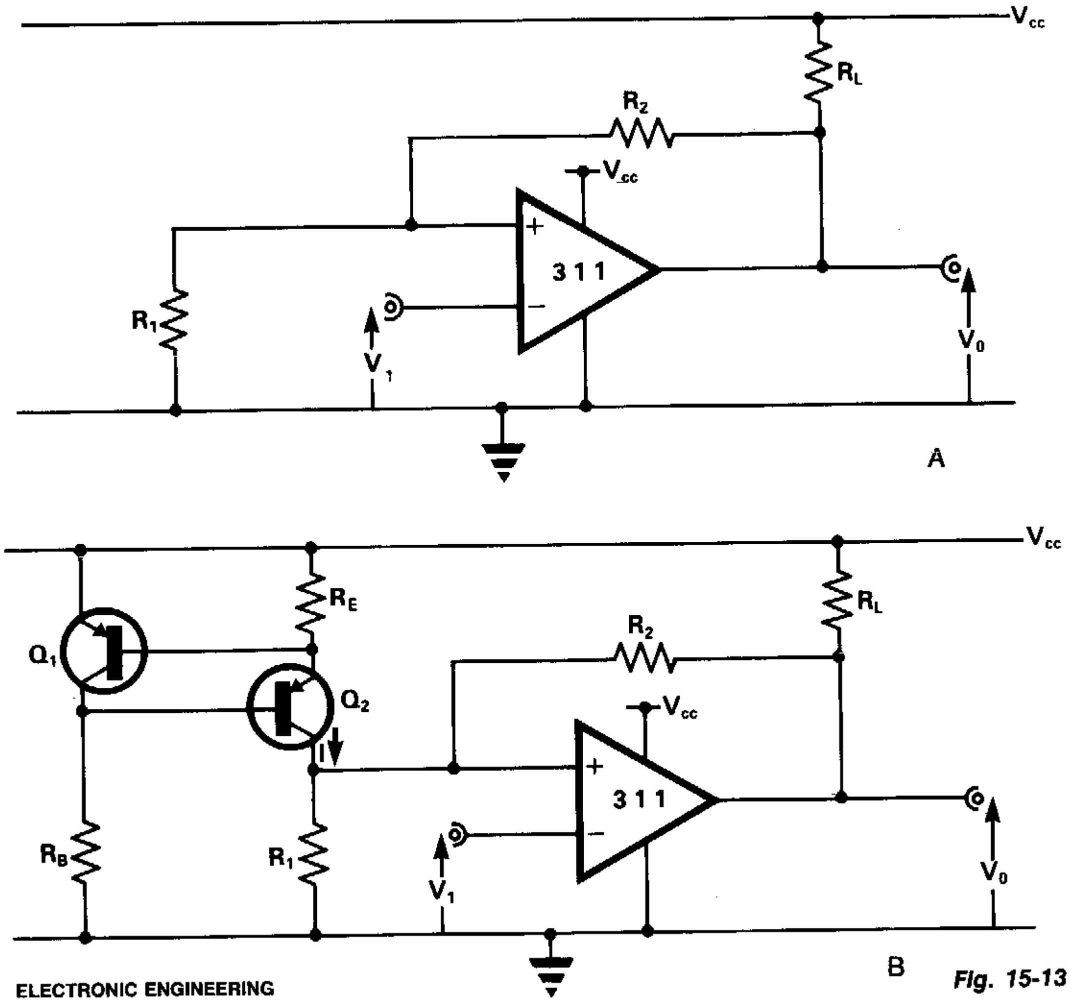
Fig. 15-12

Circuit Notes

The circuit provides unambiguous LED + or − bar readout with steps of 0.1%. The reference frequency is multiplied by the PLLIC1 and divider IC9 to output 64 × F (ref) and this is then gated by dividing F (measure) by 32 in IC8 thus is F (ref) = (measure) then IC2 counts 1024 pulses. Should the count be more than 1031 than the latch IC4c/IC4a is set to indicate count too high (F (measure) F (ref)) and if the count is less than 1017 then IC3/IC4b indicate count too low (F (measure) F (ref). These signals are latched by IC5 at the end of each period by the latch signal from IC6e.

When the two frequencies are within + or − 0.6% the LSB's of the counter IC2 are decoded and latched by IC7 and displayed on LED's IC6c resets the counter after latching the data.

TTL-COMPATIBLE SCHMITT TRIGGER



ELECTRONIC ENGINEERING

Fig. 15-13

Circuit Notes

The comparator has an output pull-up resistor R_L and is connected up to operate as a Schmitt trigger using the single rail supply V_{CC}. The feedback resistors R1 and R2 give upper and lower threshold levels V_{T+}, VVV_{TW}, respectively. V_{T+} is easily set by suitable resistor selection but there is little independent choice of V_{T-} because V_{T-} cannot exceed $V_{CE(SAT)}$. In Fig. 15-13B current-source, comprising the transistors R_E, R_B produces a current $I \sim$: (V_{EB}/R_E), V_{EB} ($\sim$ 0.65 V) being the emitter-base voltage of Q1 and Q2. Fig. 15-13C shows the results of a practical test using the circuit of Fig. 15-13B, and the following operating and component data:

V_{CC} = 5 V; R_L = 1 K ohm; R_1 = R_2 = 10 K ohm;
R_B = 3.6 K ohm; R_E = 1 K ohm + 10 K ohm pot;
Q_1 = ZTX500; Q_2 = ZTX500.

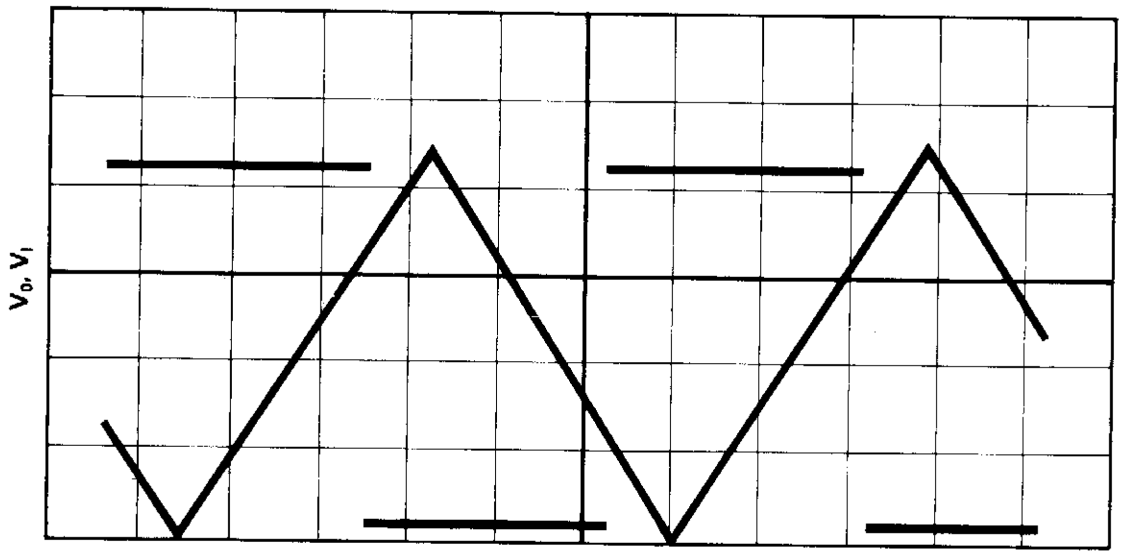


Fig. 15-13 Continued

C

16

Computer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-BIT μP BUS INTERFACE

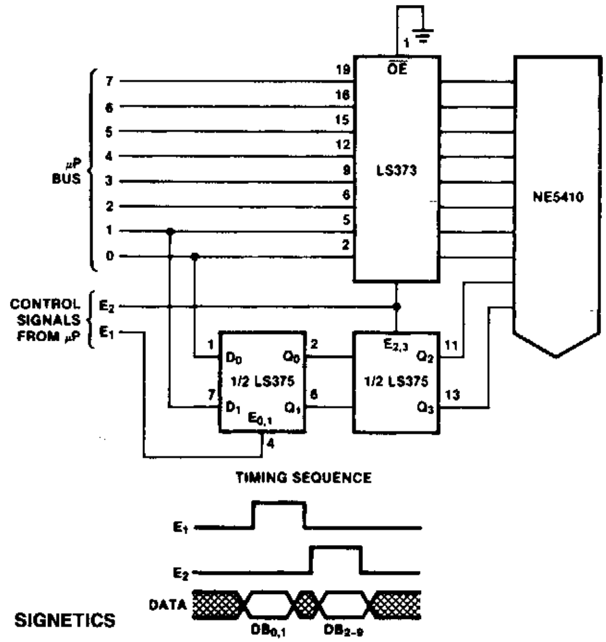


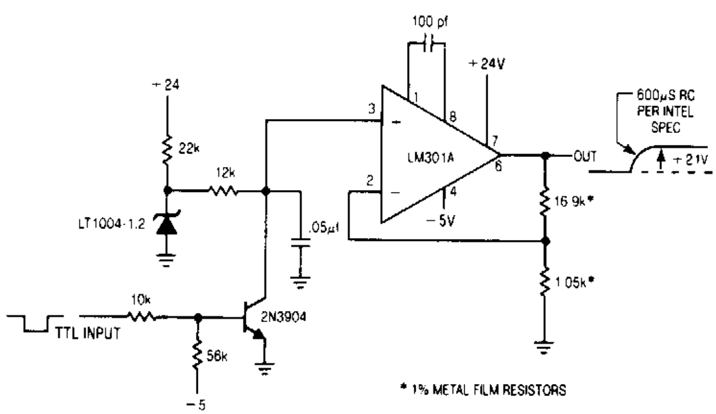
Fig. 16-1

SIGNETICS

Circuit Notes

With this double latch technique, valid data will be latched to the DAC until updated with the E_2 pulse. Timing will depend on the processor used.

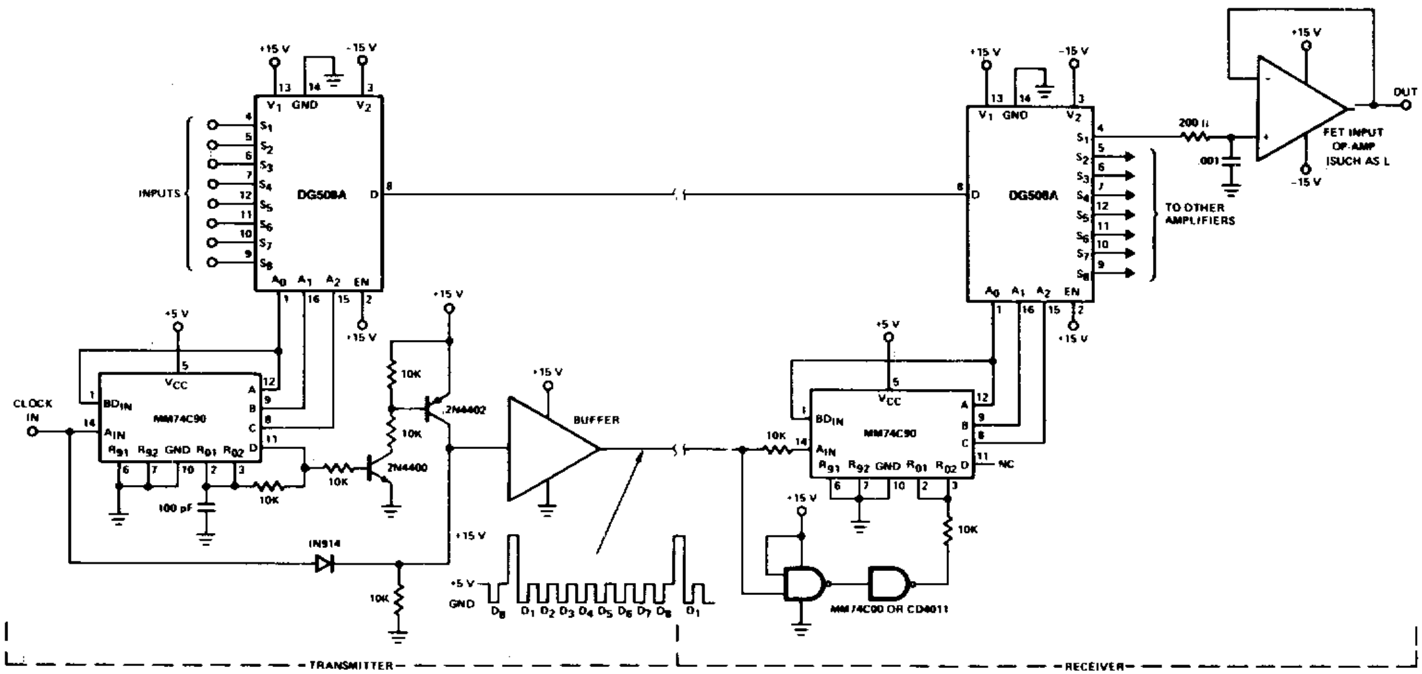
V_{PP} GENERATOR FOR EPROMS



LINEAR TECHNOLOGY CORP.

Fig. 16-2

EIGHT-CHANNEL MUX/DEMUX SYSTEM



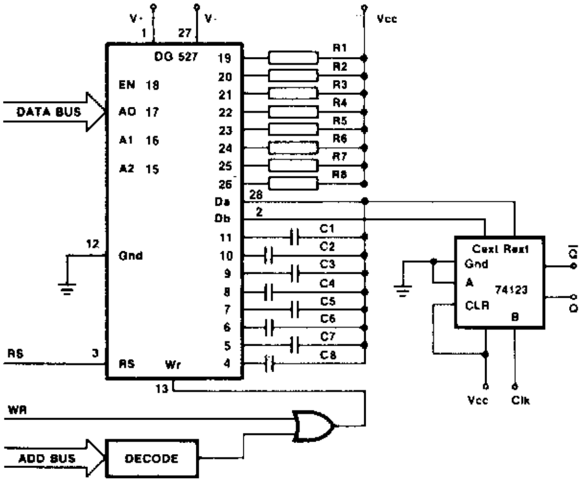
SILICONIX

Fig. 16-3

Circuit Notes

A number of signals may be sent between two points simultaneously by making a slight modification in the receiver circuit. A second DG508A is used as a demultiplexer, allowing all 8 channels to be monitored continuously.

MICROPROCESSOR SELECTED PULSE WIDTH CONTROL



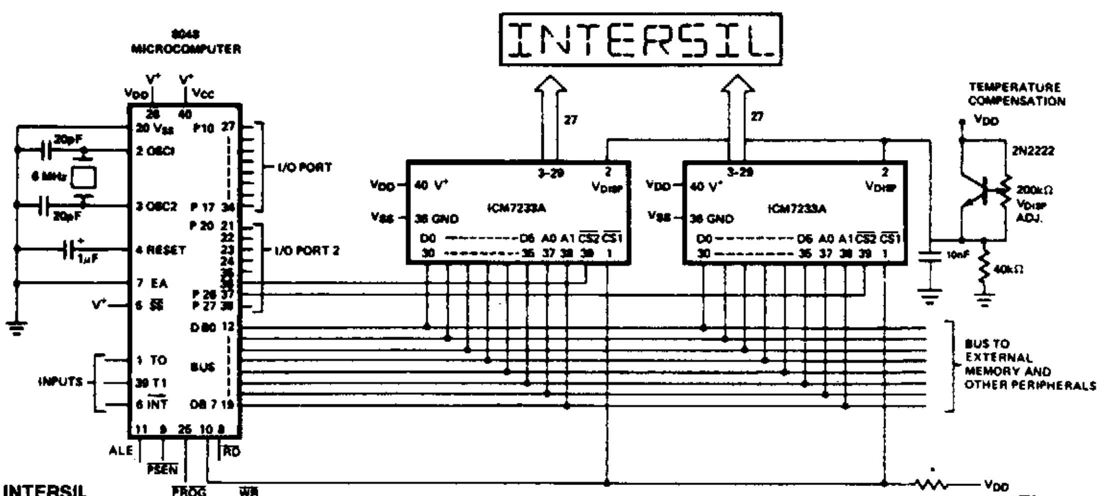
SILICONIX

Fig. 16-4

Circuit Notes

Differential multiplexers are generally used in process control applications to eliminate errors due to common mode signals. In this circuit however, advantage is taken of the dual multiplexing capability of the switch. This is achieved by using the multiplexer to select pairs of RC networks to control the pulse width of the multivibrator. This can be a particularly useful feature in process control applications where there is a requirement for a variable width sample "window" for different control signals.

8048/IM80C48 MICROCOMPUTER WITH 8-CHARACTER 16-SEGMENT ASCII TRIPLEX LIQUID CRYSTAL DISPLAY



INTERSIL

Fig. 16-5

Circuit Notes

The two bit character address is merged with the data and written to the display driver under the control of the WR line. Port lines are used to either select the target driver, or deselect all of them for other bus operations.

CMOS DATA ACQUISITION SYSTEM

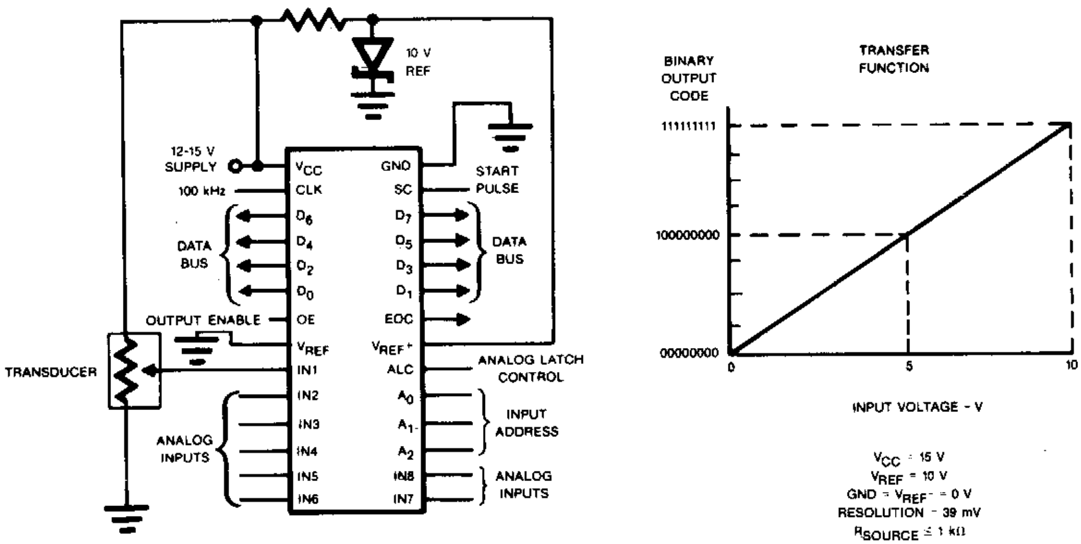


Fig. 16-6

SILICONIX

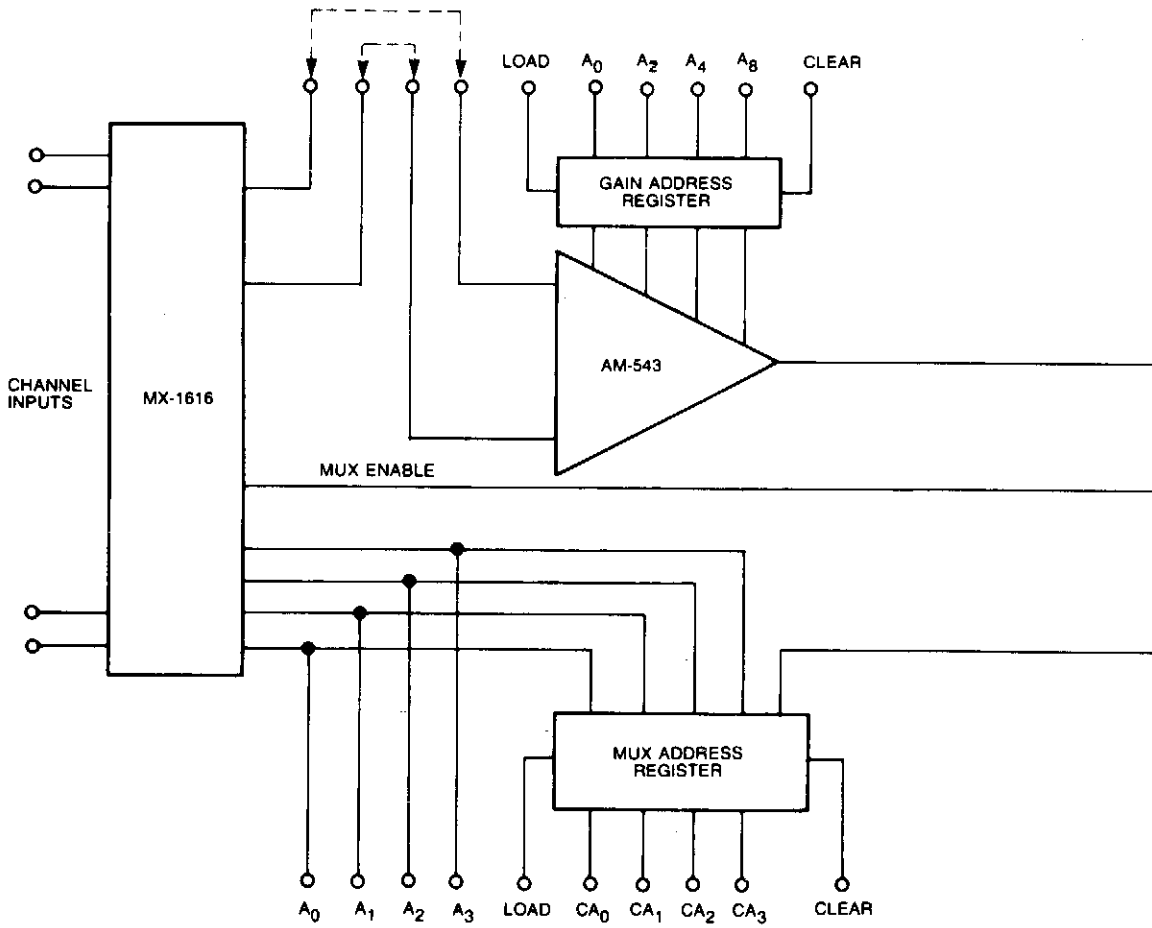
Circuit Notes

Charge redistribution to achieve A/D conversion. In typical applications, as a ratiometric conversion system for a microprocessor, V_{REF-} will be connected to ground and V_{REF+} will be connected to V_{CC}. The output will then be a simple proportional ratio between analog input voltage and V_{CC}. The general relationship is:

$$\frac{D_{OUT}}{2^8} = \frac{V_{IN}}{V_{REF+} - V_{REF-}}$$

Where D_{OUT} = Digital Output
V_{IN} = Analog Input
V_{REF} = Positive Reference Potential
V_{REF} = Negative Reference Potential

HIGH SPEED DATA ACQUISITION SYSTEM

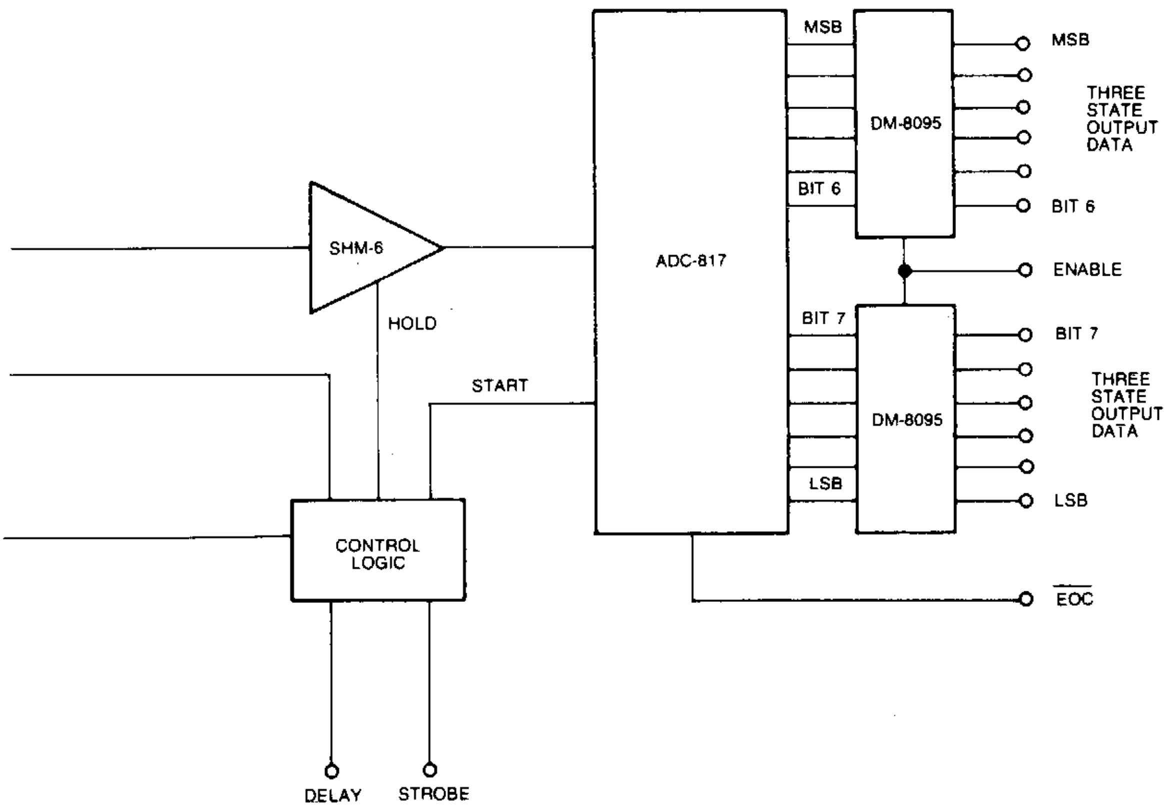


DATEL

Fig. 16-7

Circuit Notes

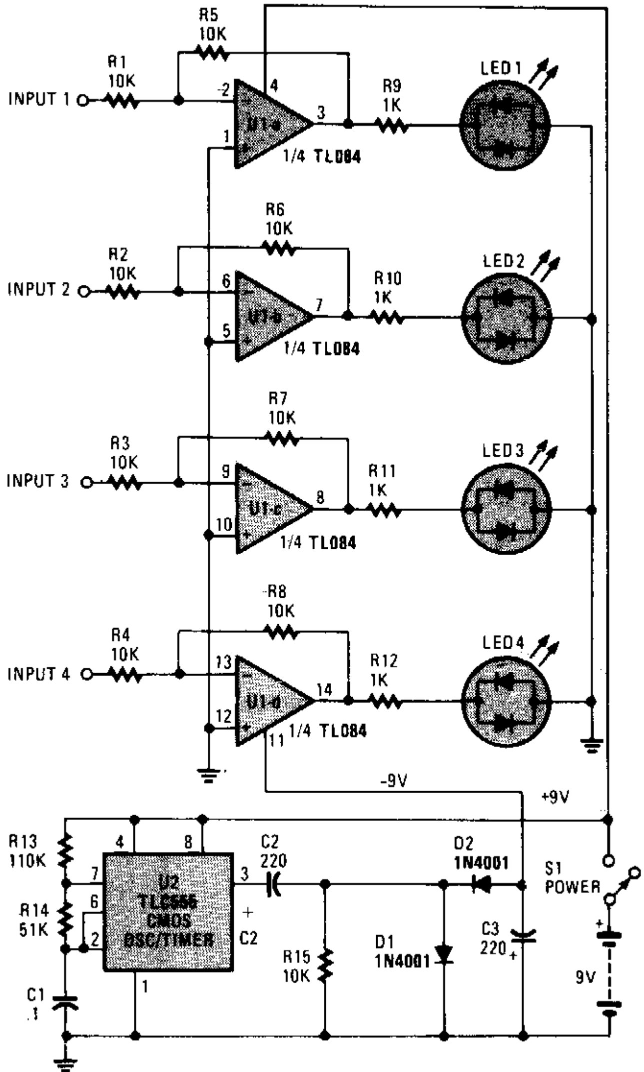
This diagram shows a high-speed data acquisition system with 8 differential inputs and 12-bit resolution using the AM-543. If the control logic is timed so that the Sample-Hold-ADC section is converting one analog value while the mux-amplifier section is allowed to settle to the next input value, throughout rates greater than 156 KHz can be achieved. The AM-543 is used with Datel's ADV-817, a 12-bit hybrid A/D with a 2 μsec conversion rate, the SHM-6, a 0.01%, 1 μsec hybrid Sample-Hold, and the MX-1616,



a low cost, high-speed monolithic analog multiplexer. The system works as follows:

The μP selects a channel and initiates a conversion at G = 1 and then looks at the MSB of the conversion result. If the MSB = 1, the μP will store the value. If the MSB = O, the μP will select G = 2. The μP will repeat the cycle of gain incrementing, comparison, and analog-to-digital conversion until the MSB = 1. The μP will then test for an output of all 1's, as this is the full-scale output of the A/D. If the output is all 1's, the μP will decrement the gain by 1 step and perform the final conversion.

BUFFERED BREAKOUT BOX



HANDS-ON ELECTRONICS

Fig. 16-8

Circuit Notes

The monitoring circuit consists of four tri-color LEDs driven by an equal number of op amps configured as gain-of-one inverting amplifiers. Each LED is wired in the circuit so that it glows red when the input to the op amp is high, and green when the input is low. The LED remains off when the input is disconnected from a circuit, when it's at ground potential, and when it's connected to a 3-state output that's in the high-impedance state. Each input has an impedance of 10,000 ohms preventing the circuit

BUFFERED BREAKOUT BOX, Continued.

from loading communication lines. The op amp requires both positive and negative supply voltages to properly drive the LEDs. Both voltages are supplied by a single, nine-volt battery. The battery supplies the positive source directly. The negative source is supplied via a CMOS 555 oscillator/timer that's configured as an astable oscillator, which is used to drive a standard diode/capacitor voltage doubler. When the 555 is connected to the monitoring circuit, the output voltage is not 18 volts (2 × 9), but a little under nine volts, due to loading. The circuit draws about 16 mA with all LEDs off; with all four on, it draws between 20 and 30 mA, depending on how many LEDs are high, and how many are low. The use of CMOS op amps reduces quiescent current drain considerably.

Z80 CLOCK

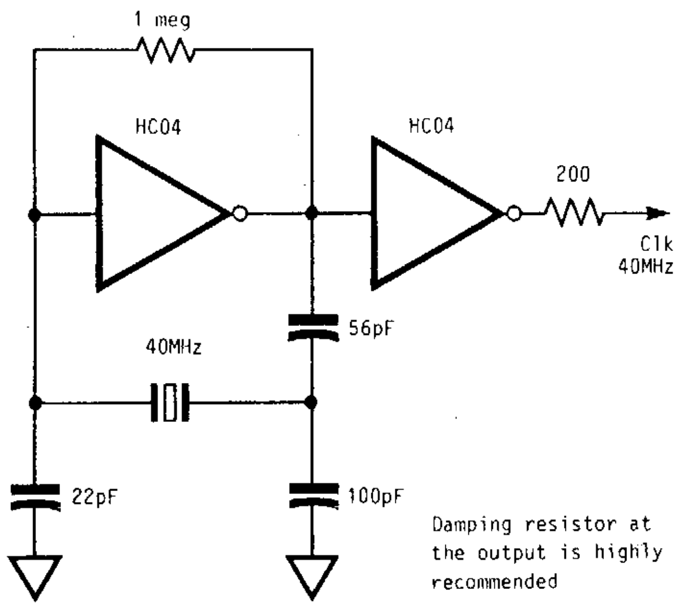


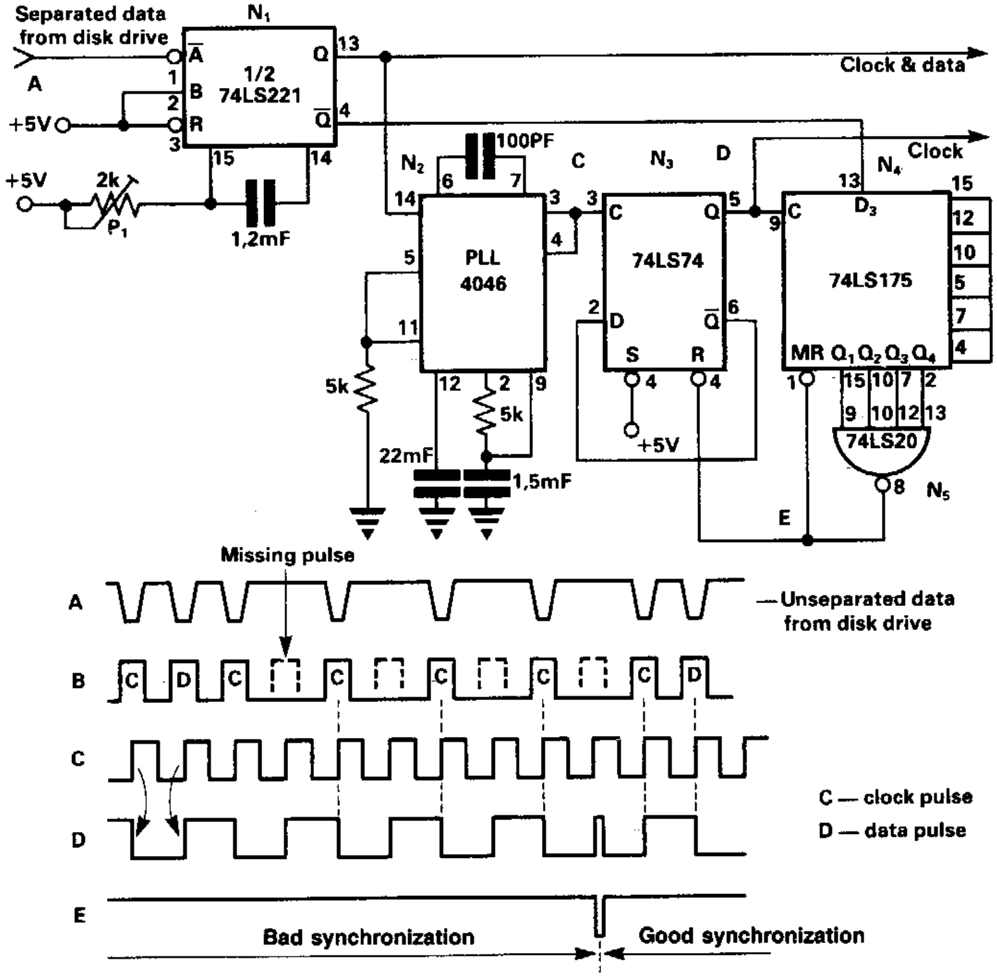
Fig. 16-9

ELECTRONIC ENGINEERING

Circuit Notes

The circuit will operate reliably from below 1 MHz to above 400 MHz. With V_{CC} = 5 V the output of the second inverter essentially attains a full swing from 0 V to 5 V. Such large logic output levels and broad frequency range capabilities make this oscillator quite suitable for driving MOS components such as CPU, controller chip, peripheral devices, as well as other TTL products. A damping resistor in series between the clock output of the oscillator and the input of the device being driven will remove the undesirable undershoot and ringing caused by the high speed CMOS part.

DATA SEPARATOR FOR FLOPPY DISKS



ELECTRONIC ENGINEERING

Fig. 16-10

Circuit Notes

The data separator is intended for use with 8″ flexible diskettes with IBM 3870 soft sectored format. The circuit delivers data and clock (B) and clock pulses (D). These two signals must be in such a sequence that the negative edge of the clock pulse is at the middle of a data cell.

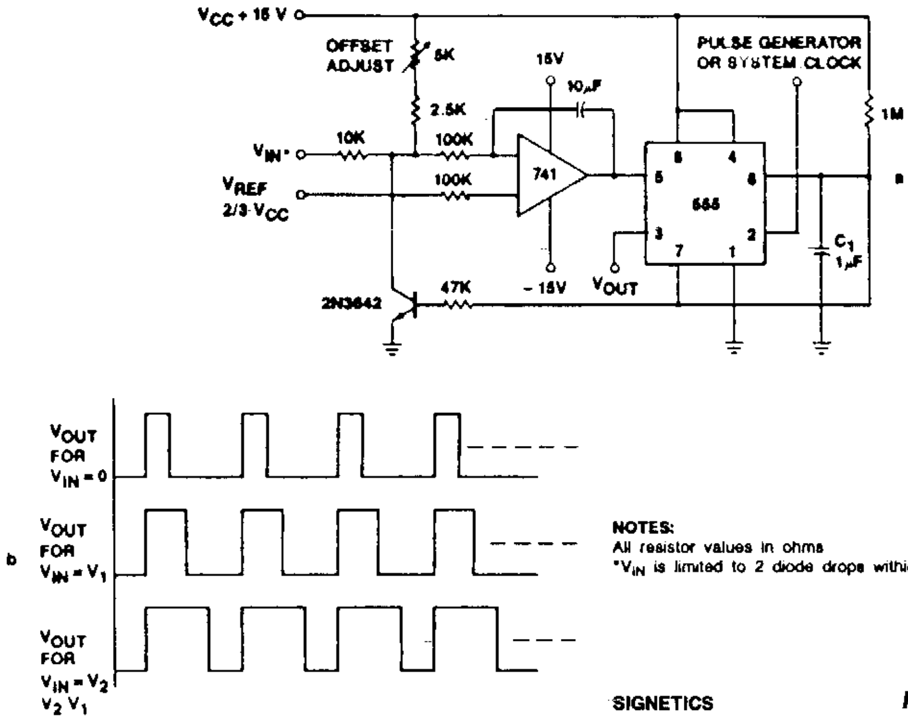
Unseparated data (A) from the floppy unit is shaped with one shot N1. Trimmer P1 should be adjusted so that pulses (B) are 1 μs wide. This signal synchronizes PLL N2 with a free running frequency adjusted to 500 kHz. The output of the PLL is 90° out of phase with its input. D-type flip-flop N3 is connected as a divider by two and changes state at each positive edge of (C). N4, connected as a shift register, looks for four consecutive missing pulses. When this happens, the circuit is resynchronized with (E) so that the negative edge of (D) is in the middle of a data cell.

17

Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VOLTAGE-TO-PULSE DURATION CONVERTER



NOTES:
All resistor values in ohms
*V_{IN} is limited to 2 diode drops within ground or below V_{CC}

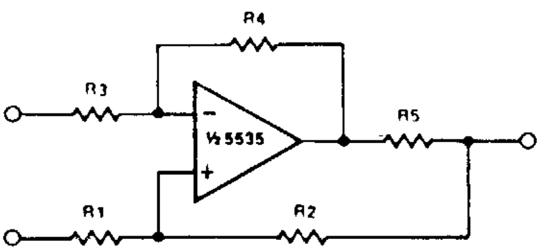
SIGNETICS

Fig. 17-1.

Circuit Notes

Voltage levels can be converted to pulse durations by combining an op amp and a timer IC. Accuracies to better than 1% can be obtained with this circuit (a), and the output signals (b) still retain the original frequency, independent of the input voltage.

VOLTAGE-TO-CURRENT CONVERTERS



NOTES:

$$\frac{R_2}{R_1} = \frac{R_4}{R_3}$$

$$I_{OUT} = \frac{-V_{IN} \cdot R_2}{R_5 \cdot R_1}$$

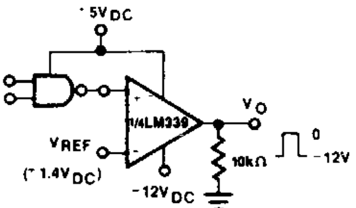
SIGNETICS

Circuit Notes

A simple voltage-to-current converter is shown in the figure. The current out is I_{OUT} or V_{IN}/R. For negative currents, a pnp can be used and, for better accuracy, a Darlington pair can be substituted for the transistor. With careful design, this circuit can be used to control currents of many amps. Unity gain compensation is necessary.

Fig. 17-2

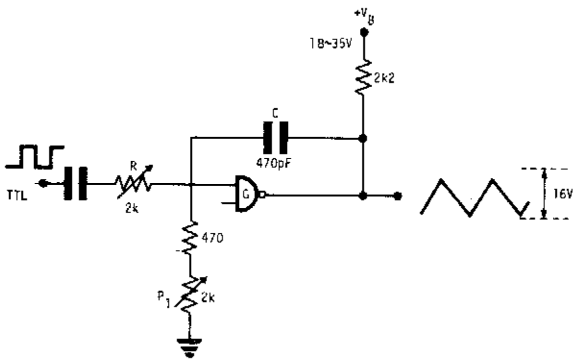
TTL-TO-MOS LOGIC CONVERTER



SIGNETICS

Fig. 17-3

TTL SQUARE WAVE-TO-TRIANGLE CONVERTER



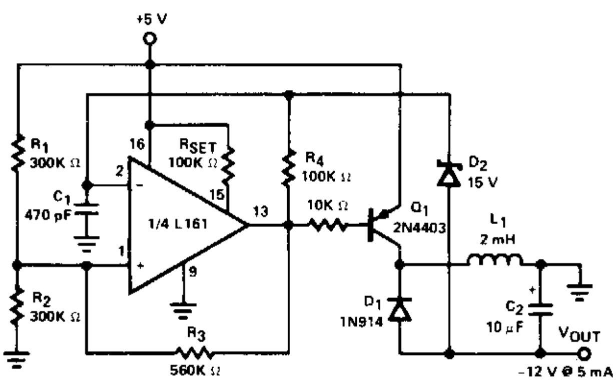
ELECTRONIC ENGINEERING

Fig. 17-4

Circuit Notes

This fixed frequency triangular waveform generator driven by a TTL square wave generates typically 16-V p-p triangles at frequencies up to several MHz. It uses only one NAND open collector gate, or one open collector inverter as a fast integrator with gain. Careful successive adjustments of R and P1 are needed. When correct adjustments are reached, output amplitude and linearity are largely independent of the value of V_B, from a minimum of 18 V up to 35 V. The value of C shown is for 100 kHz; at higher frequencies, it must be reduced in proportion.

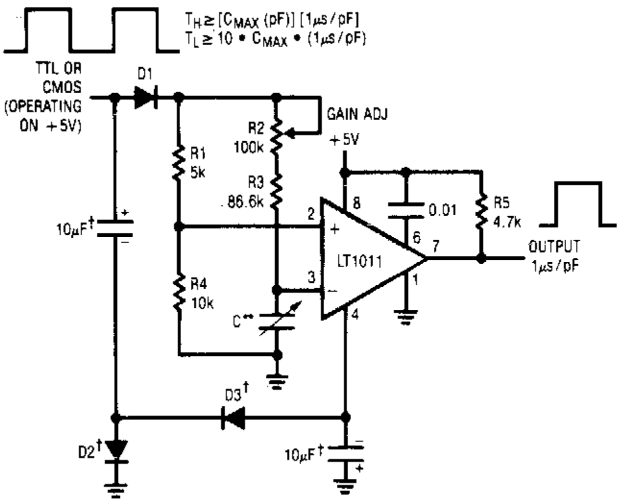
A REGULATED DC-TO-DC CONVERTER



SILICONIX

Fig. 17-5

CAPACITANCE TO PULSE WIDTH CONVERTER



$$*PW = (R2 + R3)(C)\left(\frac{R1 + R4}{R1}\right)$$, INPUT CAPACITANCE OF LT1011 IS ≈ 6pF. THIS IS AN OFFSET TERM.

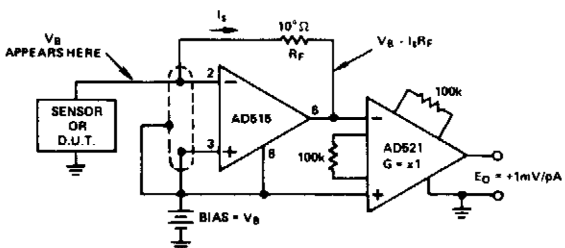
† THESE COMPONENTS MAY BE ELIMINATED IF NEGATIVE SUPPLY IS AVAILABLE (−1V TO −15V).

** TYPICAL 2 SECTIONS OF 365pF VARIABLE CAPACITOR WHEN USED AS SHAFT ANGLE INDICATION.

LINEAR TECHNOLOGY CORP.

Fig. 17-6

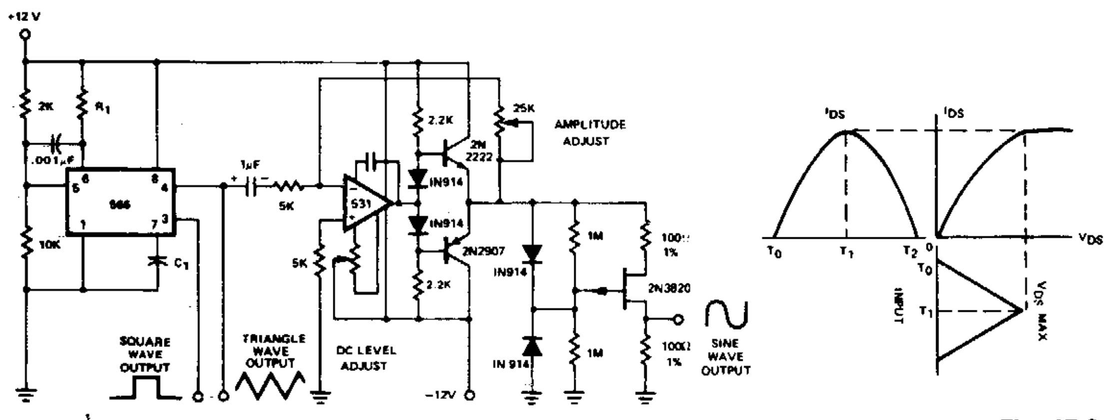
CURRENT-TO-VOLTAGE CONVERTER WITH GROUNDED BIAS AND SENSOR



ANALOG DEVICES, INC.

Fig. 17-7

TRIANGLE-TO-SINE CONVERTERS



$f_o \approx \frac{1}{3R_1C_1}$

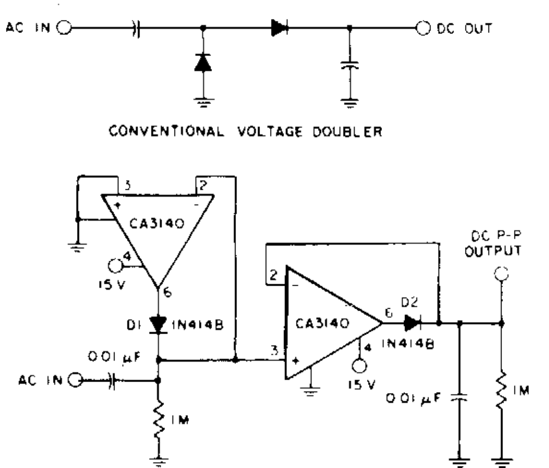
SIGNETICS

Fig. 17-8

Circuit Notes

Conversion of triangle wave shapes to sinusoids is usually accomplished by diode-resistor shaping networks, which accurately reconstruct the sine wave segment by segment. Two simpler and less costly methods may be used to shape the triangle waveform of the 566 into a sinusoid with less than 2% distortion. The non-linear $I_{DS}V_{DS}$ transfer characteristic of a P-channel junction FET is used to shape the triangle waveform. The amplitude of the triangle waveform is critical and must be carefully adjusted to achieve a low distortion sinusoidal output. Naturally, where additional waveform accuracy is needed, the diode-resistor shaping scheme can be applied to the 566 with excellent results since it has very good output amplitude stability when operated from a regulated supply.

PRECISION PEAK-TO-PEAK AC-DC CONVERTER



Circuit Notes

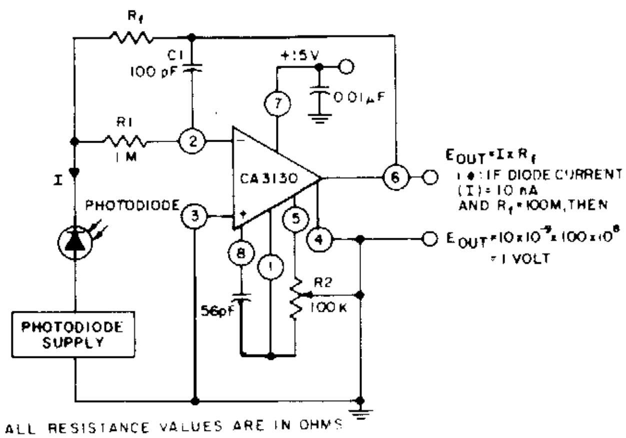
Using a CA3140 BiMOS op amp and a single positive supply converts a conventional voltage doubler with two precision diodes into a precision peak-to-peak ac-to-dc voltage converter having wide dynamic range and wide bandwidth.

ALL RESISTANCE VALUES ARE IN OHMS

GENERAL ELECTRIC/RCA

Fig. 17-9

PHOTODIODE CURRENT-TO-VOLTAGE CONVERTER



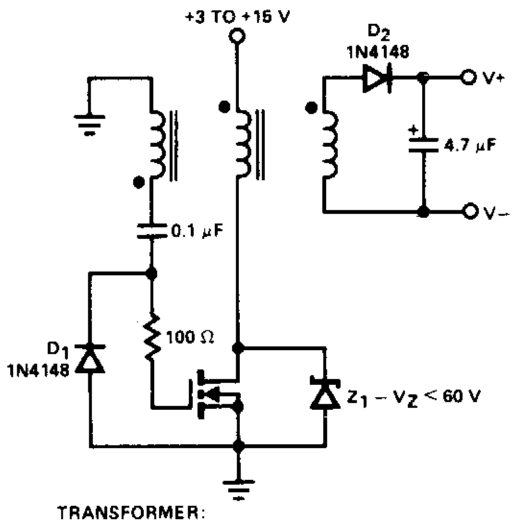
GENERAL ELECTRIC/RCA

Fig. 17-10

Circuit Notes

The circuit uses three CA3130 BiMOS op amps in an application sensitive to sub-picoampere input currents. The circuit provides a ground-referenced output voltage proportional to input current flowing through the photodiode.

SELF OSCILLATING FLYBACK CONVERTER



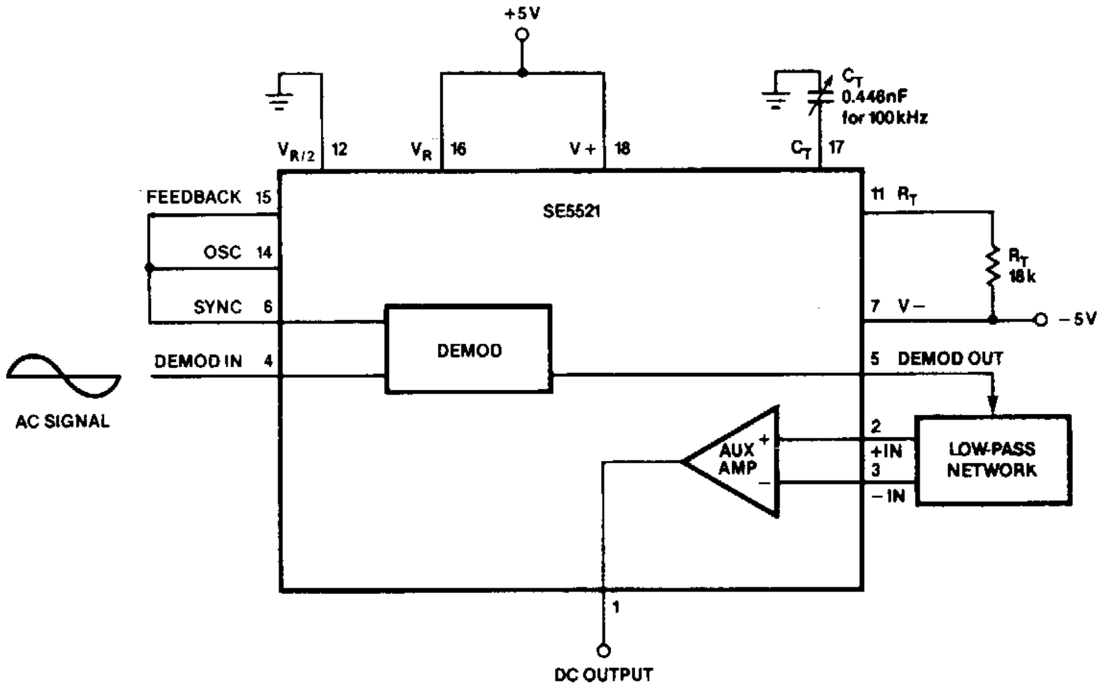
Circuit Notes

Low-power converter uses the core characteristics to determine frequency. With the transformer shown, operating frequency is 250 kHz. Diode D1 prevents negative spikes from occurring at the MOSFET gate, the 100 ohm resistor is a parasitic suppressor, and Z1 serves as a dissipative voltage regulator for the output and also clips the drain voltage to a level below the rated power FET breakdown voltage.

SILICONIX, INC.

Fig. 17-11

RMS-TO-DC CONVERTER



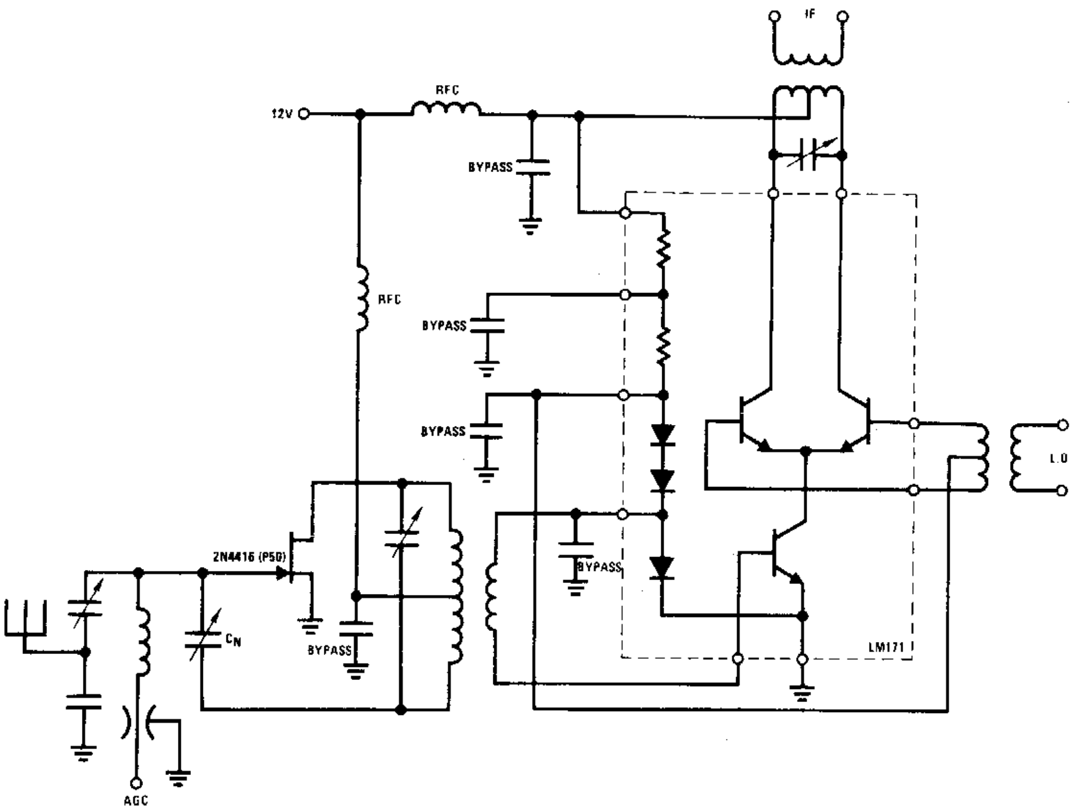
NOTE:
1. The DC output at Pin 1 varies linearly with the RMS input at Pin 4.
2. C_T is tweaked until the sync signal is in phase with the AC signal.

SIGNETICS *Fig. 17-12*

Circuit Notes

An ac voltmeter may be easily constructed. Simplicity of the circuit and low component count make it particularly attractive. The demodulator output is a full-wave rectified signal from the ac input at Pin 4. The dc component on the rectified signal at Pin 5 varies linearity with the rms input at Pin 4 and thus provides an accurate rms-to-dc conversion at the output of the filter (Pin 1). C_T is a variable capacitor that is tweaked until the oscillator signal to the sync input of the demodulator is in phase with the ac signal at Pin 4.

100 MHz CONVERTER

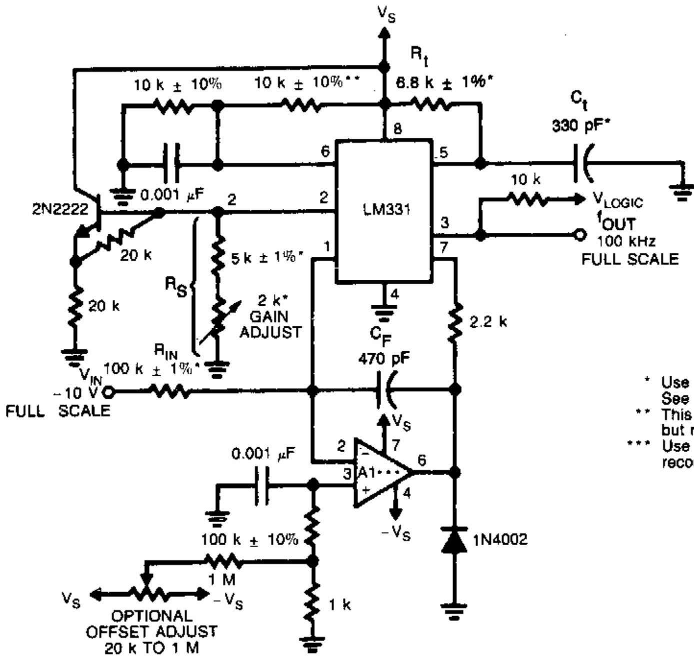


NATIONAL SEMICONDUCTOR CORP. *Fig. 17-13*

Circuit Notes

The 2N4416 JFET will provide noise figures of less than 3 dB and power gain of greater than 20 dB. The JFET's outstanding low crossmodulation and low intermodulation distortion provides an ideal characteristic for an input stage. The output feeds into an LM171 used as a balanced mixer. This configuration greatly reduces local oscillator radiation both into the antenna and into the *if* strip and also reduces *rf* signal feedthrough.

PRECISION VOLTAGE-TO-FREQUENCY CONVERTER



* Use stable components with low temperature coefficients. See Typical Applications section.
** This resistor can be 5 kΩ or 10 kΩ for V_S = 8 V to 22 V, but must be 10 kΩ = 4.5 V to 8 V.
*** Use low offset voltage and low offset current op amps for A1: recommended types LF411A or LF356.

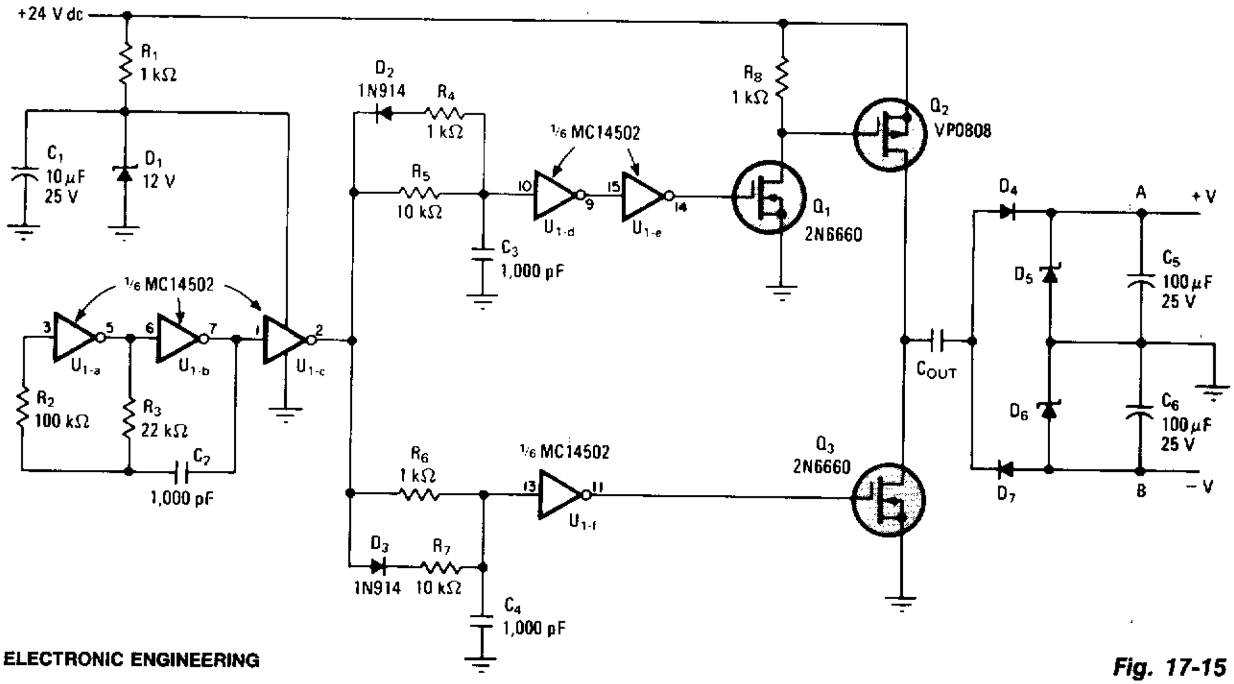
NATIONAL SEMICONDUCTOR CORP.

Fig. 17-14

Circuit Notes

In this circuit, integration is performed by using a conventional operational amplifier and feedback capacitor, C_F. When the integrator's output crosses the nominal threshold level at pin 6 of the LM131, the timing cycle is initiated. The average current fed into the op amp's summing point (pin 2) is i × (1.1 R_tC_t) × f which is perfectly balanced with $-V_{IN}/R_{IN}$. In this circuit, the voltage offset of the LM131 input comparator does not affect the offset or accuracy of the V-to-F converter as it does in the stand-alone V-to-F converter, nor does the LM131 bias current or offset current. Instead, the offset voltage and offset current of the operational amplifier are the only limits on how small the signal can be accurately converted.

BIPOLAR DC-DC CONVERTER REQUIRES NO INDUCTOR



ELECTRONIC ENGINEERING

Fig. 17-15

Circuit Notes

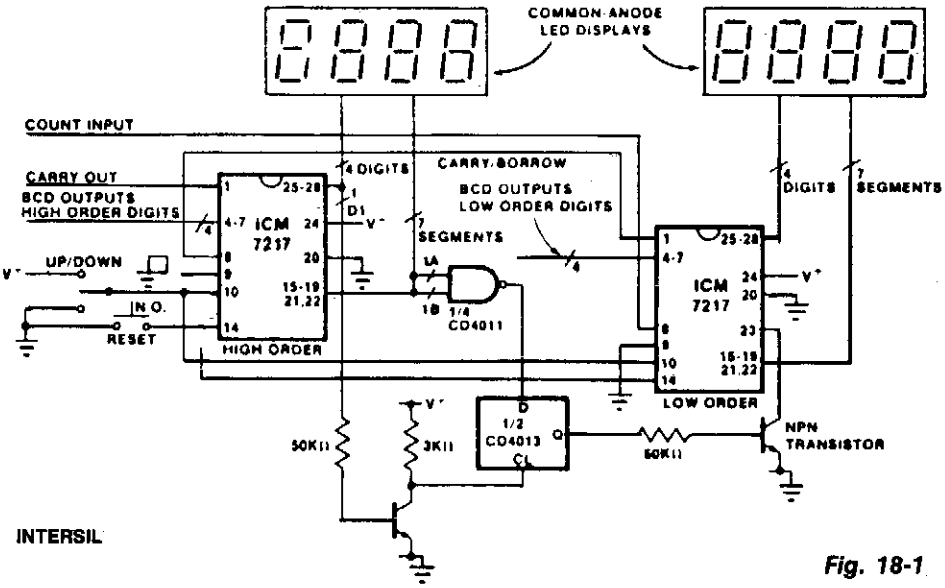
Inverters U1a and U1b form a 20-kilohertz oscillator whose square wave output—further shaped by D2, R4, and R5 and by D3, R6, and R7—drives power field-effect transistors Q2 and Q3. The p-channel and n-channel FETs conduct alternately, in a push-pull configuration. When Q2 conducts, the positive charge on C_{out} forces diode D4 to conduct as well, which produces a positive voltage, determined by zener diode D5, at terminal A. Similarly, when Q3, in its turn conducts, the negative charge on C_{out} forces D7 to do so as well. A negative voltage, therefore, develops at terminal B, whose level is set by D6.

18

Counters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-DIGIT UP/DOWN COUNTER



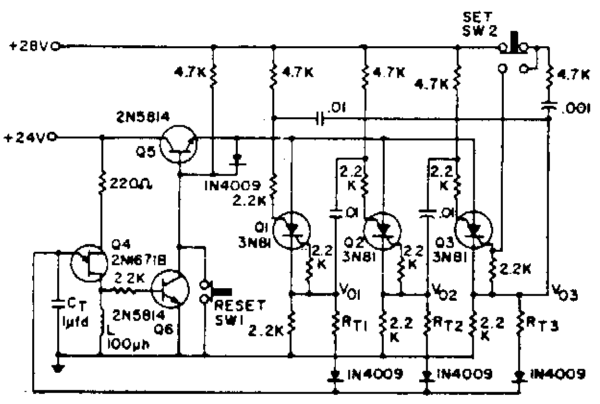
INTERSIL

Fig. 18-1

Circuit Notes

This circuit shows how to cascade counters and retain correct leading zero blanking. The NAND gate detects whether a digit is active since one of the two segments a or b is active on any unblanked number. The flip flop is clocked by the least significant digit of the high order counter, and if this digit is not blanked, the Q output of the flip flop goes high and turns on the npn transistor, thereby inhibiting leading zero blanking on the low order counter.

RING COUNTER WITH VARIABLE TIMING



GENERAL ELECTRIC

Fig. 18-2

Circuit Notes

Shift pulses are generated by the unijunction transistors. The intervals between pulses are controlled by C_T and R_T. A different R_T can be selected for each stage of the counter as shown.

20 kHz RING COUNTER

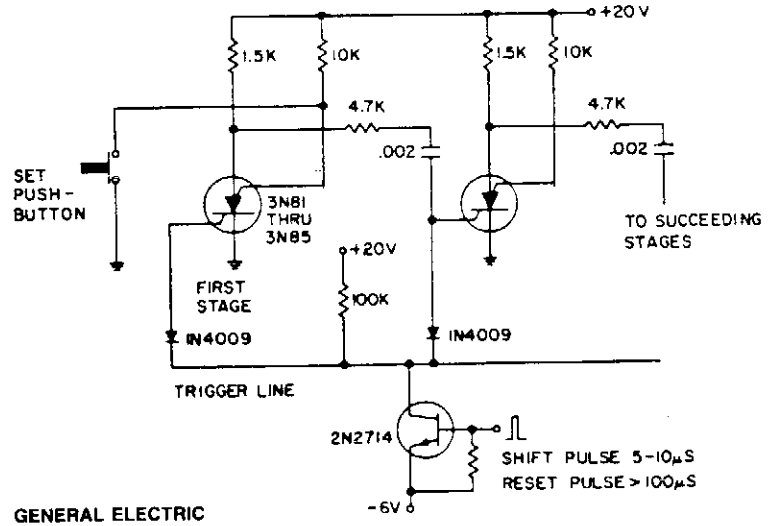


Fig. 18-3

Circuit Notes

The shift pulse turns off the conducting scs by reverse biasing the cathode gate. The charge stored on the coupling capacitor then triggers the next stage. An excessively long shift pulse charges up all the capacitors, turning off all stages. Grounding an anode gate will "set" that stage.

BINARY COUNTER

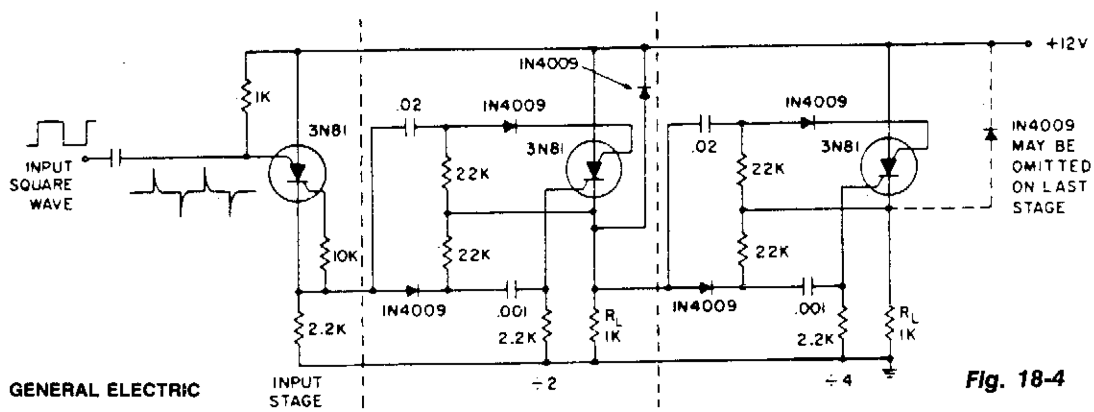
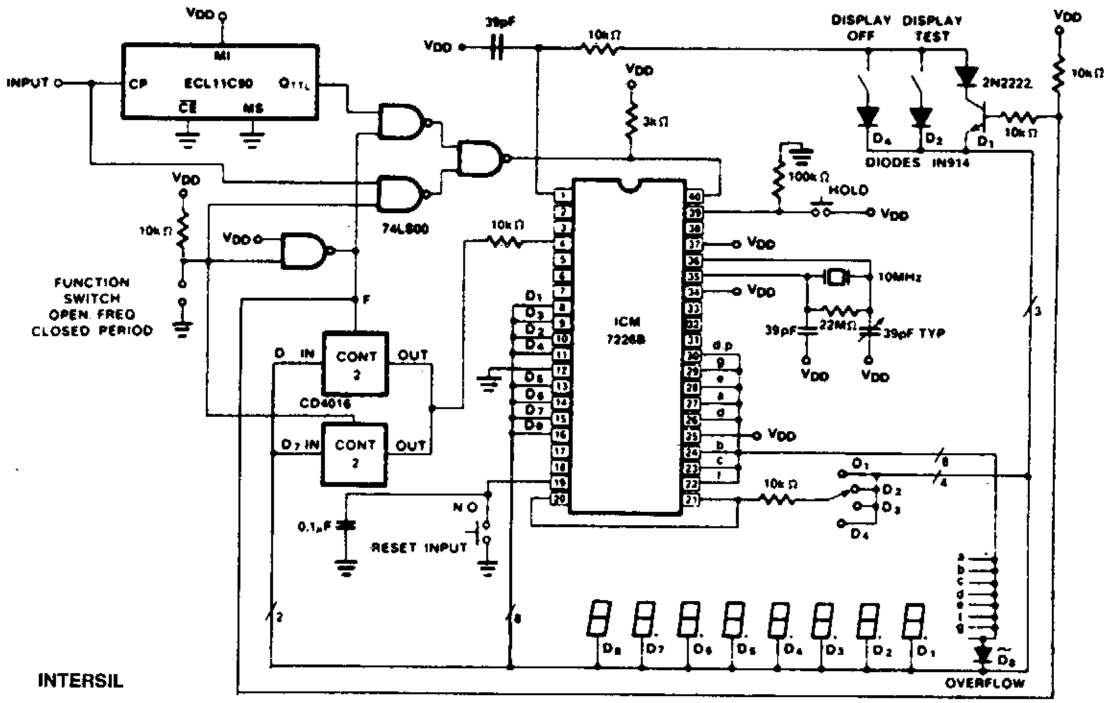


Fig. 18-4

Circuit Notes

Stages are triggered by the positive going edge. The scs is turned on at the cathode gate; turned off at the anode gate. The anode-to-cathode IN4009 suppresses positive transients while the scs is recovering. The input stage generates fast positive edges to trigger the counter.

100 MHz FREQUENCY, PERIOD COUNTER



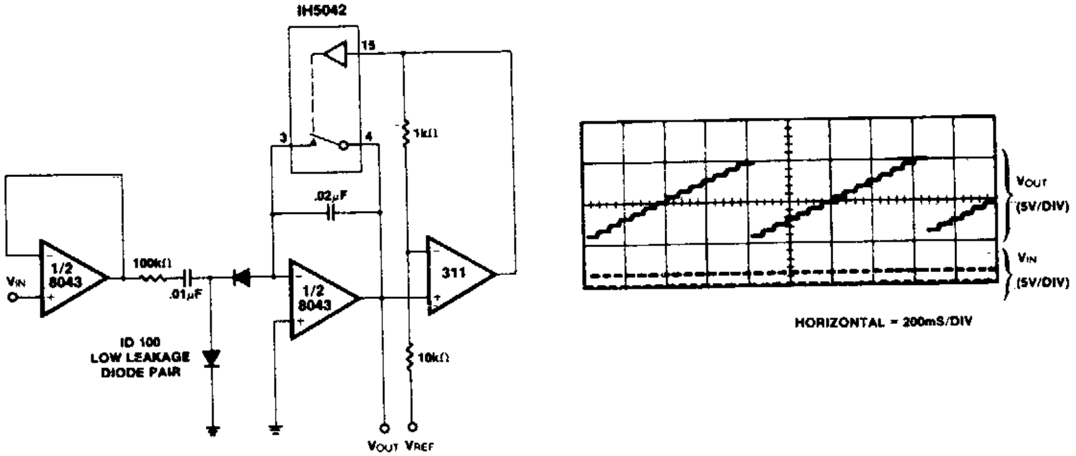
INTERSIL

Fig. 18-5

Circuit Notes

The figure shows the use of a CD4016 analog multiplex to multiplex the digital outputs back to the FUNCTION input. Since the CD4016 is a digitally controlled analog transmission gate, no level shifting of the digit output is required. The CD4051's or CD4052's could also be used to select the proper inputs for the multiplexed input on the ICM7226 from 2 or 3 bit digital inputs. These analog multiplexers may also be used in systems in which the mode of operation is controlled by a microprocessor rather than directly from front panel switches. TTL multiplexers such as the 74LS153 or 74LS251 may also be used, but some additional circuitry will be required to convert the digit output to TTL compatible logic levels.

ANALOG COUNTER CIRCUIT



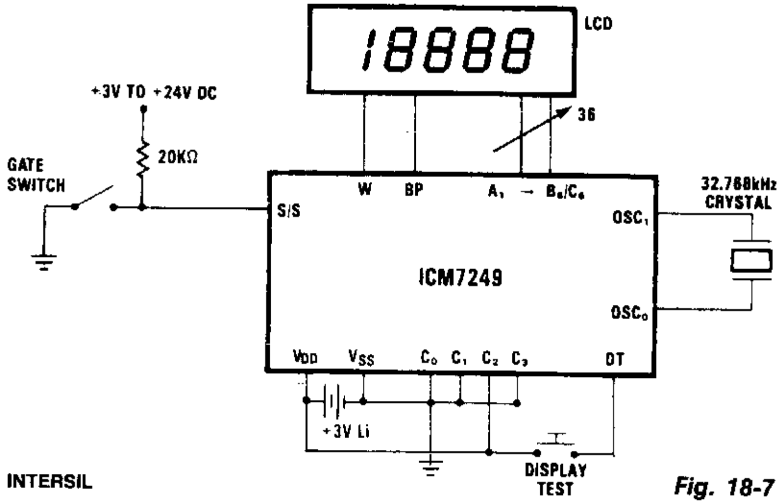
INTERSIL

Fig. 18-6

Circuit Notes

A straightforward circuit using a LM311 for the level detector and a CMOS analog gate to discharge the capacitor is shown. An important property of this type of counter is the ease with which the count can be changed; it is only necessary to change the voltage at which the comparator trips. A low cost A-D converter can also be designed using the same principle since the digital count between reset periods is directly proportional to the analog voltage used as a reference for the comparator. A considerable amount of hysteresis is used in the comparator. This ensures that the capacitor is completely discharged during the reset period. In a more sophisticated circuit, a dual comparator ''window detector'' could be used, the lower trip point is set close to ground to ensure complete discharge. The upper trip point could then be adjusted independently to determine the pulse count.

ATTENDANCE COUNTER



INTERSIL

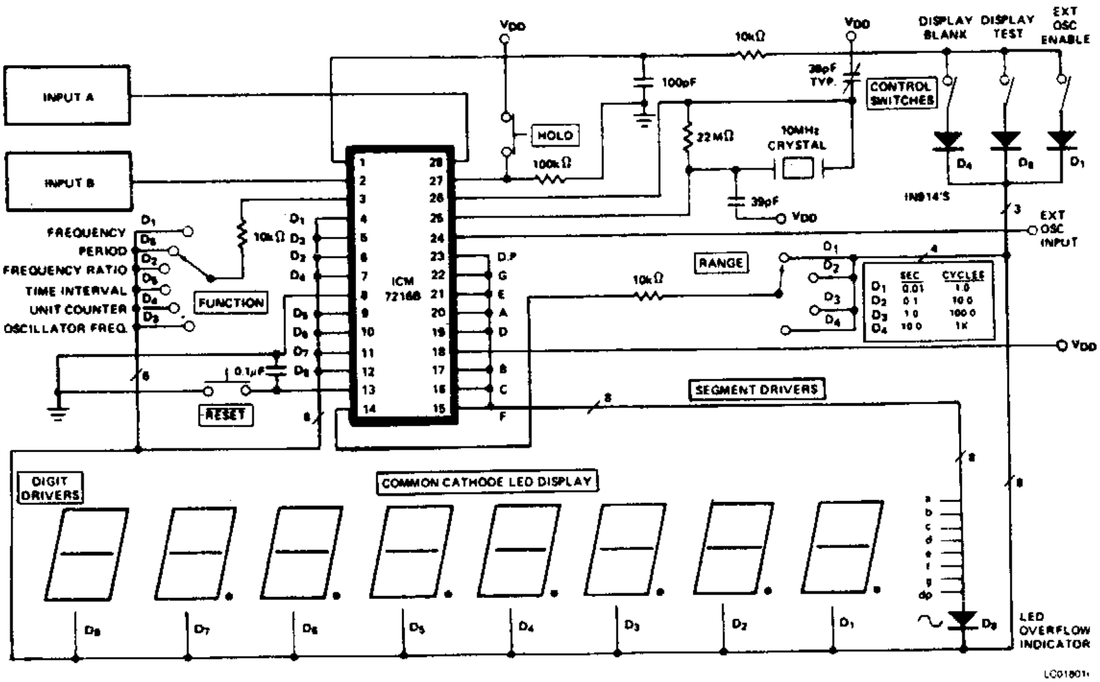
Fig. 18-7

Circuit Notes

The display shows each increment. By using mode 2, external debouncing of the gate switch is unnecessary, provided the switch bounce is less than 35ms. The 3 V lithium battery can be replaced without disturbing operation if a suitable capacitor is connected in parallel with it. The display should be disconnected, if possible, during the procedure to minimize current drain. The capacitor should be large enough to store charge for the amount of time needed to physically replace the battery (t = VC/1). A 100 μF capacitor initially charged to 3 V will supply a current of 1.0 μA for 50 seconds before its voltage drops to 2.5 V, which is the minimum operating voltage for the ICM7249.

Before the battery is removed, the capacitor should be placed in parallel, across the V_{DD} and GND terminals. After the battery is replaced, the capacitor can be removed and the display reconnected.

10 MHz UNIVERSAL COUNTER



INTERSIL

Fig. 18-8

Circuit Notes

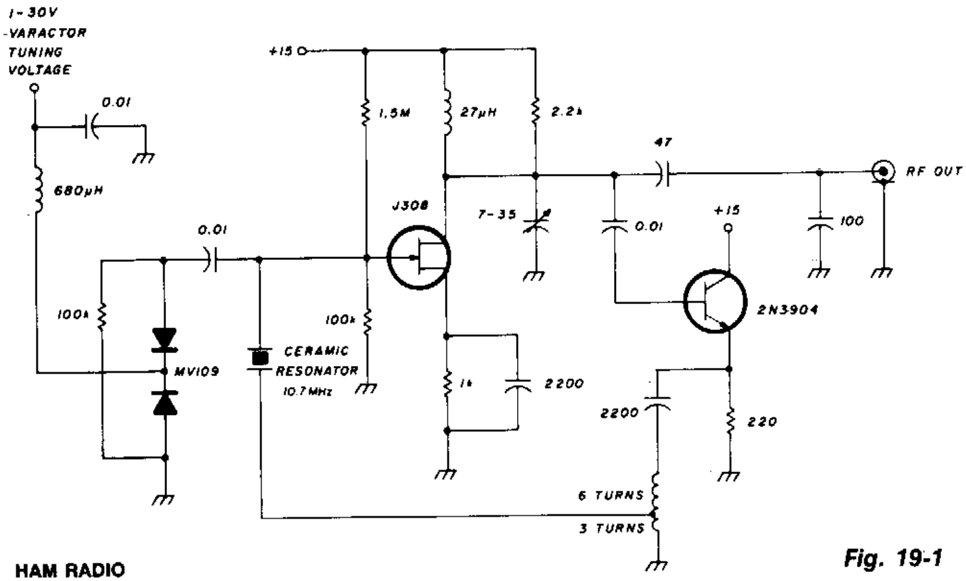
The ICM7216A or B can be used as a minimum component complete Universal Counter. This circuit can use input frequencies up to 10 MHz at INPUT A and 2 MHz at INPUT B. If the signal at INPUT A has a very low duty cycle it may be necessary to use a 74121 monostable multivibrator or similar circuit to stretch the input pulse width to be able to guarantee that it is at least 50 ns in duration.

19

Crystal Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VARACTOR-TUNED 10 MHz CERAMIC RESONATOR OSCILLATOR



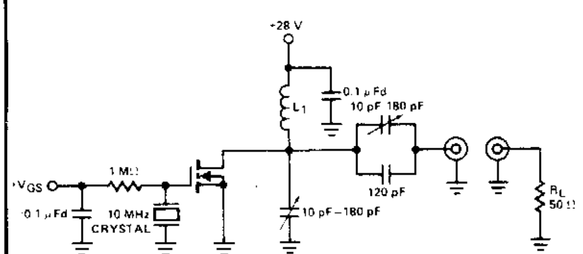
HAM RADIO

Fig. 19-1

Circuit Notes

The FET input amplifier has fixed bias with source feedback. This provides a very high input impedance with very low capacitance. The FET amplifier drives an emitter follower which, in spite of the fact that it has a low output impedance, feeds a transformer with a 3:1 turns ratio for a nine-fold impedance reduction. The result is an impedance at the ceramic resonator of a few ohms maximum. The varactor-tuned ceramic resonator oscillator has a significant frequency-temperature coefficient. The tuning range of the VCO is approximately 232 kHz, with a temperature coefficient of 350 Hz per degree centigrade. When using this circuit as a VCO, the entire 232 kHz range cannot be used because some of the tuning range must be sacrificed for the temperature dependence. If the required tuning range were 200 kHz, leaving 32 kHz for temperature variation, the resulting temperature variation would be more than 90°C.

10 MHz CRYSTAL OSCILLATOR



SILICONIX INC.

Fig. 19-2

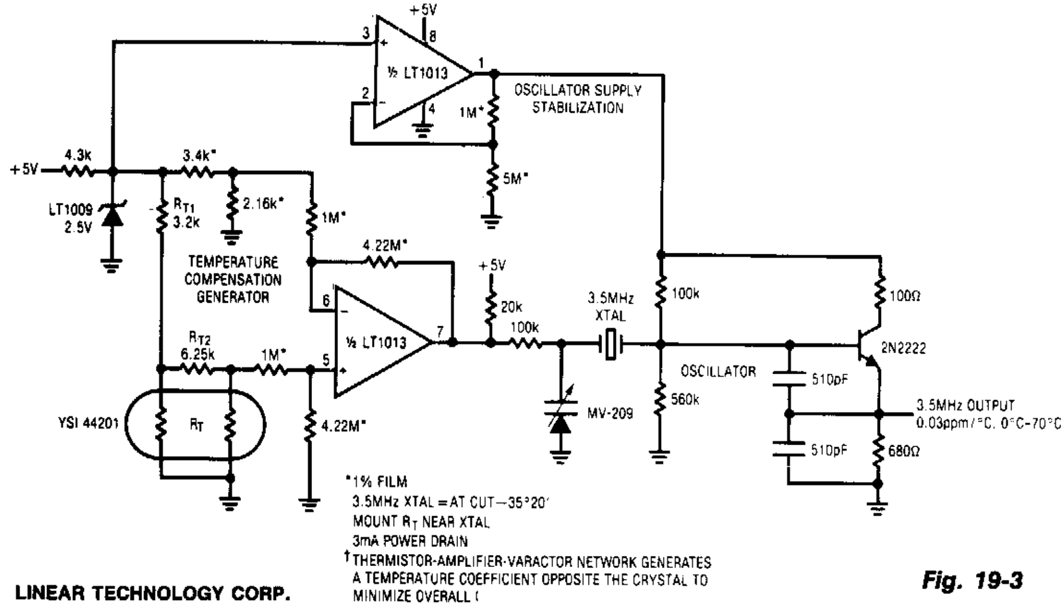
Circuit Notes

This xtal oscillator is a FET equivalent of a vacuum tube tuned to plate-tuned grid xtal oscillator. Feedback is via the drain to gate capacitance.

Parts List

L_1 ~ 18 turns #22 enameled wire on micrometals T-50-6 torroid core. ≈ 1.0 μH.

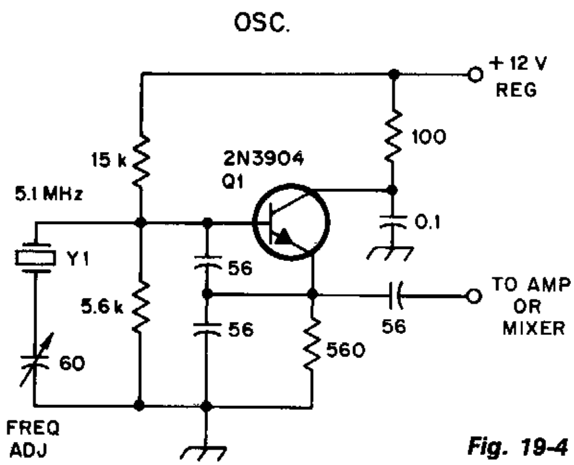
LOW POWER, 5 V DRIVEN, TEMPERATURE COMPENSATED CRYSTAL OSCILLATOR (TXCO)



LINEAR TECHNOLOGY CORP.

Fig. 19-3

CRYSTAL-CONTROLLED LOCAL OSCILLATOR FOR SSB TRANSMITTER



QST

Fig. 19-4

Circuit Notes

This oscillator may contain several switched crystals to provide channelized operation. A buffer amplifier may be added, if desired.

CRYSTAL OSCILLATOR

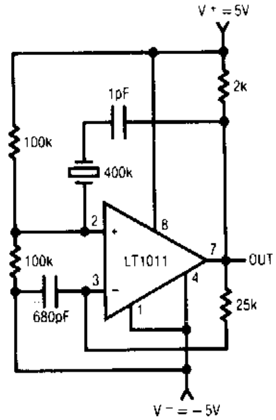


Fig. 19-5

LINEAR TECHNOLOGY CORPORATION

Circuit Notes

This circuit uses an LT1011 comparator biased in its linear mode and a crystal to establish its resonant frequency. This circuit can achieve a few hundred kHz, temperature independent clock frequency with nearly 50% duty cycle.

CRYSTAL-CONTROLLED SIGNAL SOURCE

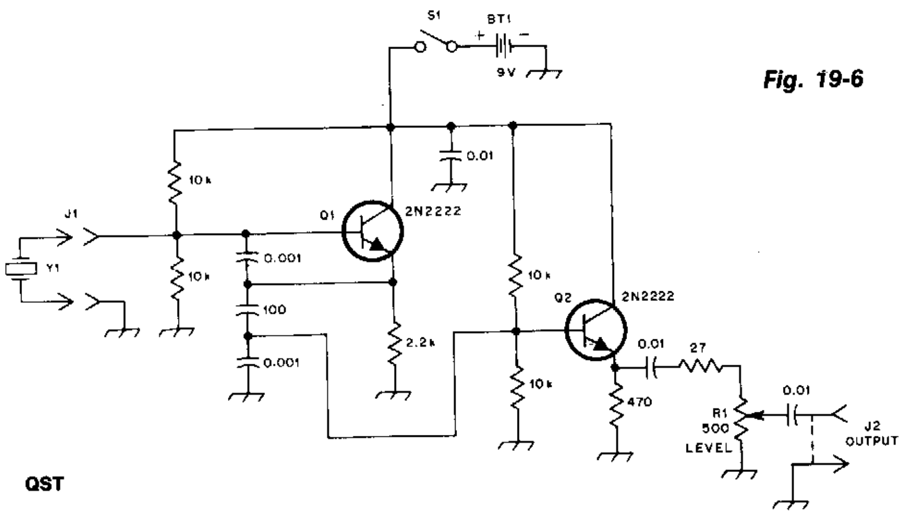


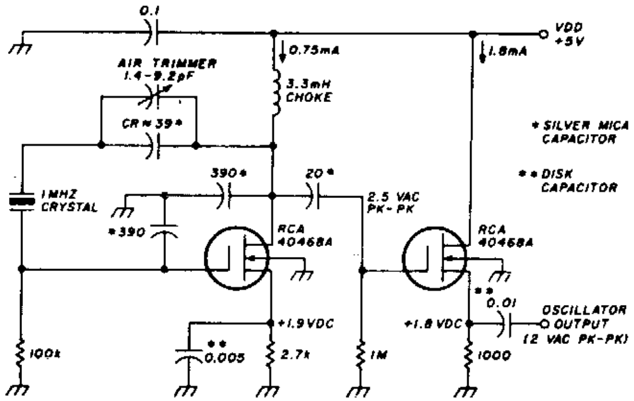
Fig. 19-6

QST

Circuit Notes

This general purpose signal source serves very well in signal-tracing applications. The output level is variable to more than 1 Vrms into a 50 Ω load. Almost any crystal in the 1 to 15 MHz range can be used. Q1 forms a Colpitts oscillator with the output taken from the emitter. A capacitive voltage divider (across the 2.2 K emitter resistor) reduces the voltage applied to the buffer amplifier, Q2. The buffer and emitter follower, provides the low input impedance necessary to drive 50 Ω loads.

1 MHz FET CRYSTAL OSCILLATOR



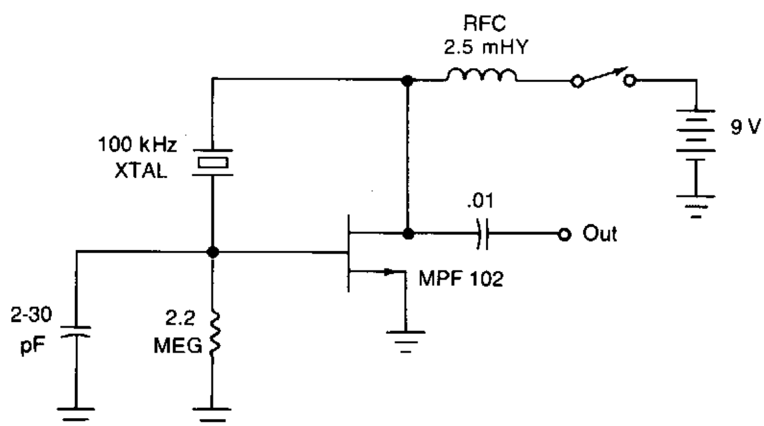
HAM RADIO

Fig. 19-7

Circuit Notes

This stable oscillator circuit exhibits less than 1 Hz frequency change over a V_{DD} range of 3–9 volts. Stability is attributed to the use of MOSFET devices and the use of stable capacitors.

PIERCE CRYSTAL OSCILLATOR



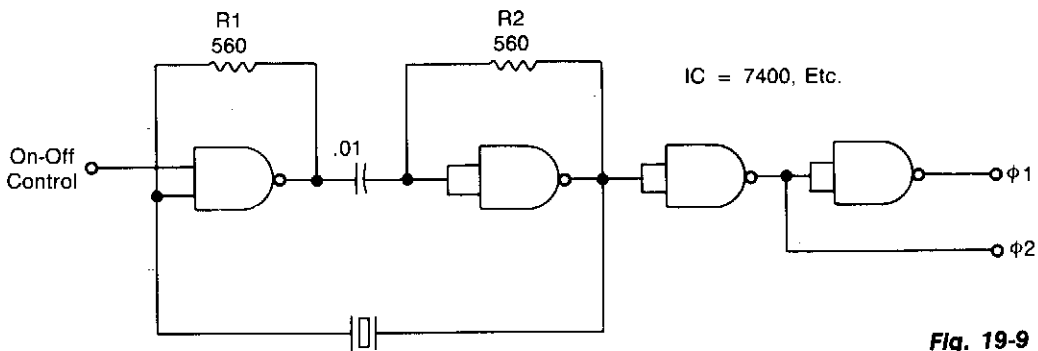
WILLIAM SHEETS

Fig. 19-8

Circuit Notes

The JFET Pierce oscillator is stable and simple. It can be the clock of a microprocessor, a digital timepiece or a calculator. With a probe at the output, it can be used as a precise injection oscillator for troubleshooting. Attach a small length of wire at the output and this circuit becomes a micropower transmitter.

IC-COMPATIBLE CRYSTAL OSCILLATOR



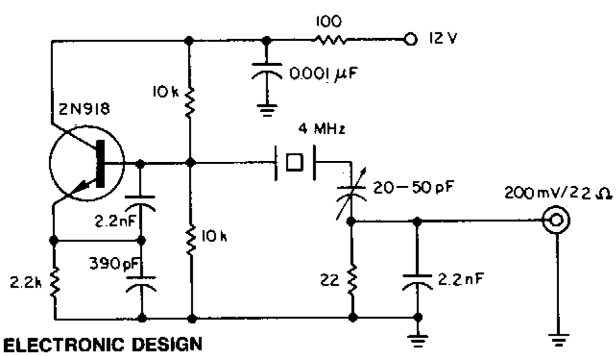
WILLIAM SHEETS

Fig. 19-9

Circuit Notes

Resistors R1 and R2 temperature-stabilize the NAND gates; they also ensure that the gates are in a linear region for starting. Capacitor C1 is a dc block; it must have less than 1/10 ohm impedance at the operating frequency. The crystal runs in a series-resonant mode. Its series resistance must be low; AT-cut crystals for the 1- to 10-MHz range work well. The output waveshape has nearly a 50% duty cycle, with chip-limited rise times. The circuit starts well from 0° to 70°C.

CRYSTAL OSCILLATOR PROVIDES LOW NOISE



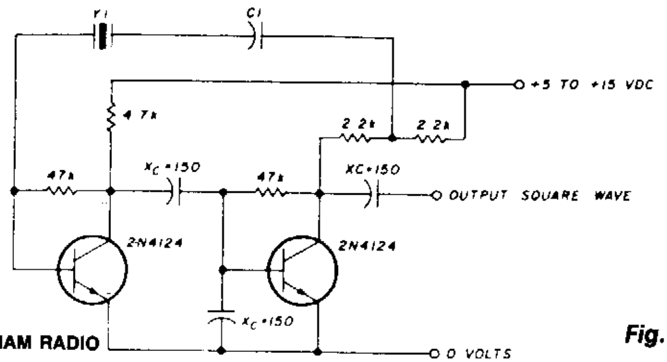
ELECTRONIC DESIGN

Fig. 19-10

Circuit Notes

The oscillator delivers an output of high spectral purity without any substantial sacrifice of the usual stability of a crystal oscillator. The crystal in addition to determining the oscillator's frequency, is used also as a low-pass filter for the unwanted harmonics and as a bandpass filter for the sideband noise. The noise bandwidth is limited to less than 100 Hz. All higher harmonics are substantially suppressed—60 dB down for the third harmonic of the 4-MHz fundamental oscillator frequency.

LOW-FREQUENCY CRYSTAL OSCILLATOR—10 kHz to 150 kHz



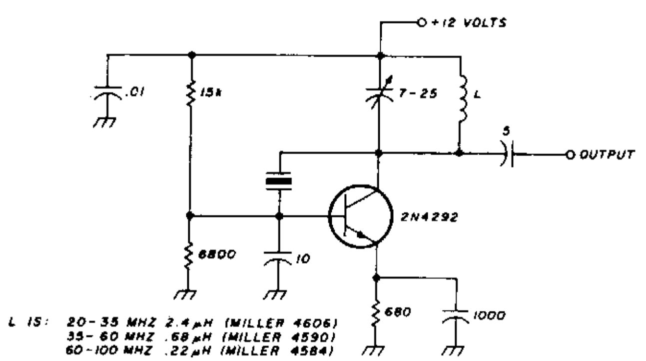
HAM RADIO

Fig. 19-11

Circuit Notes

C1 in series with the crystal may be used to adjust the oscillator output frequency. Value may range between 20 pF and 0.01 μF, or may be a trimmer capacitor and will approximately equal the crystal load capacitance. X values are approximate and can vary for most circuits and frequencies; this is also true for resistance values. Adequate power supply decoupling is required; local decoupling capacitors near the oscillator are recommended. All leads should be extremely short in high frequency circuits.

OVERTONE CRYSTAL OSCILLATOR



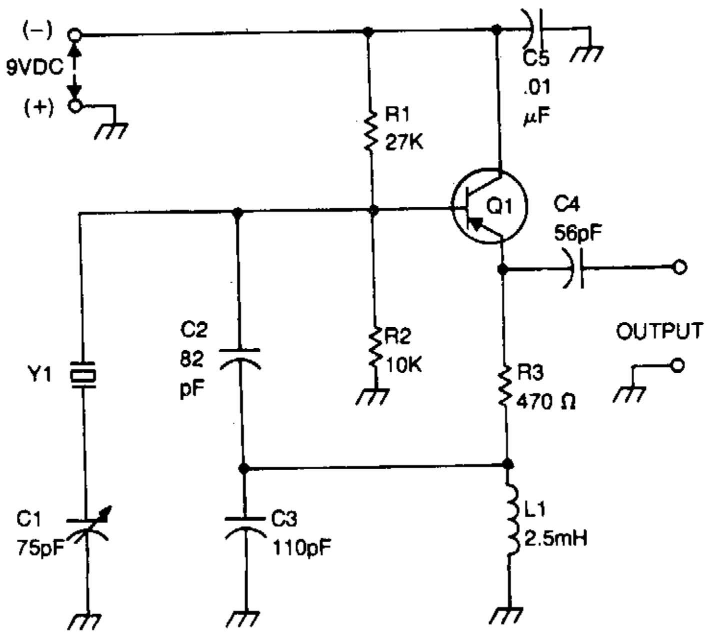
HAM RADIO

Fig. 19-12

Circuit Notes

This oscillator is designed for overtone crystals in the 20–100 MHz range operating in the third and fifth mode. Operating frequency is determined by the tuned circuit.

COLPITTS OSCILLATOR



TAB BOOKS, INC.

Fig. 19-13

Circuit Notes

Bias for the pnp bipolar transistor is provided by resistor voltage divider network R1/R2. The collector of the oscillator transistor is kept at ac ground by capacitor C5, placed close to the transistor. Feedback is provided by capacitor voltage divider C2/C3.

CRYSTAL-CONTROLLED OSCILLATOR

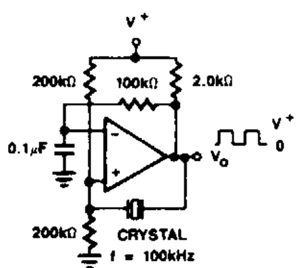
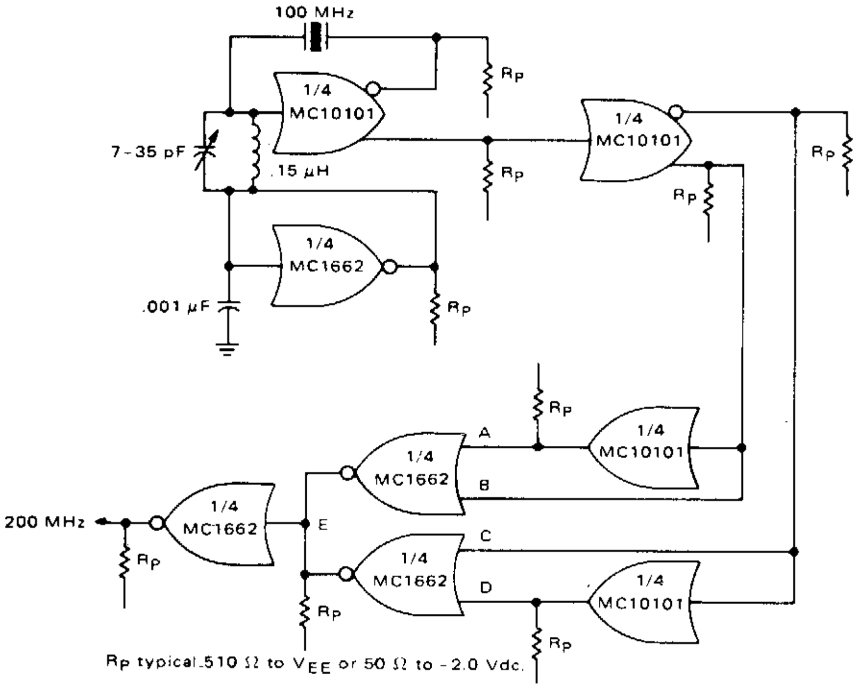


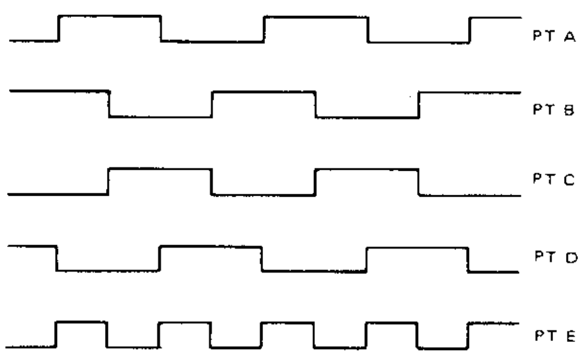
Fig. 19-14

SIGNETICS

HIGH-FREQUENCY CRYSTAL OSCILLATOR



200 MHz Crystal Oscillator



Frequency Doubler Waveforms

MOTOROLA, INC.

Fig. 19-15

HIGH-FREQUENCY CRYSTAL OSCILLATOR, Continued.

Circuit Notes

A high speed oscillator is possible by combining an MECL 10 K crystal oscillator with an MECL III frequency doubler as shown. One section of the MC10101 is connected as a 100 MHz crystal oscillator with the crystal in series with the feedback loop. The LC tank circuit tunes the 100 MHz harmonic of the crystal and may be used to calibrate the circuit to the exact frequency. A second section of the MC10101 buffers the crystal oscillator and gives complementary 100 MHz signals. The frequency doubler consists of two MC10101 gates as phase shifters and two MC1662 NOR gates. For a 50% duty cycle at the output, the delay to the true and complement 100 MHz signals should be 90°. This may be built precisely with 2.5 ns delay lines for the 200 MHz output or approximated by the two MC10101 gates. The gates are easier to incorporate and cause only a slight skew in output signal duty cycle. The MC1662 gates combine the 4 phase 100 MHz signals as shown in Figure B. The outputs of the MC1662's are wire-OR connected to give the 200 MHz signal. MECL III gates are used because of the bandwidth required for 200 MHz signals. One of the remaining MC1662 gates is used as a V_{BB} bias generator for the oscillator. By connecting the NOR output to the input, the circuit stays in the center of the logic swing or at V_{BB}. A 0.001 μF capacitor ensures the V_{BB} circuit does not oscillate.

CRYSTAL-CONTROLLED OSCILLATOR OPERATES FROM ONE MERCURY CELL

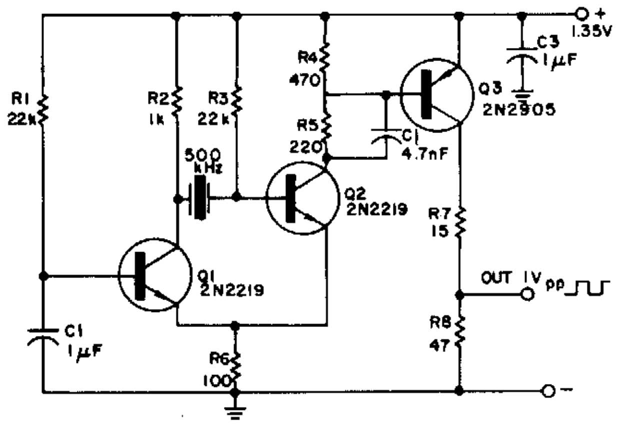


Fig. 19-16

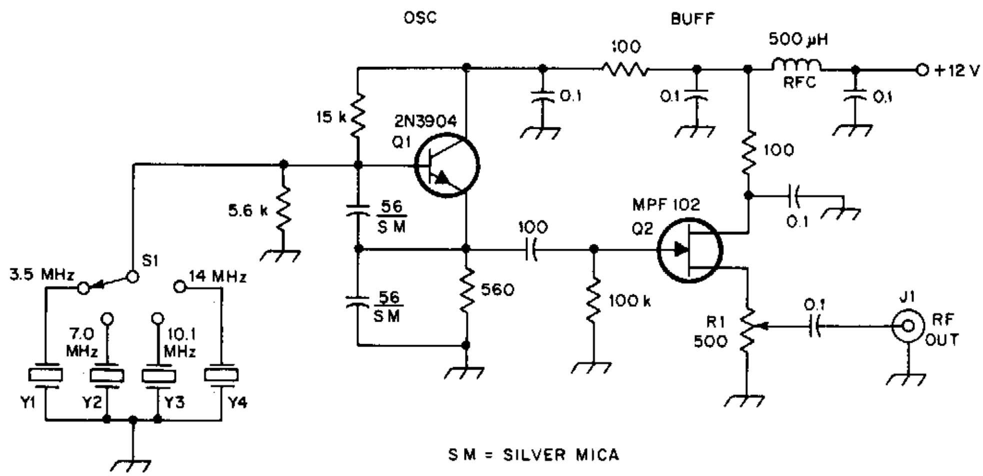
ELECTRONIC DESIGN

Inexpensive crystal controlled oscillator operates from a 1.35-volt source.

Circuit Notes

The circuit is powered by a single 1.35 V mercury cell and provides a 1 V square-wave output. As shown, the crystal is a tuned circuit between transistors Q1 and Q2, which are connected in the common-emitter configuration. Positive feedback provided by means of R permits oscillation. The signal at the collector of Q2 is squared by Q3, which switches between cutoff and saturation. R7 permits short-circuit-proof operation.

HIGH FREQUENCY SIGNAL GENERATOR



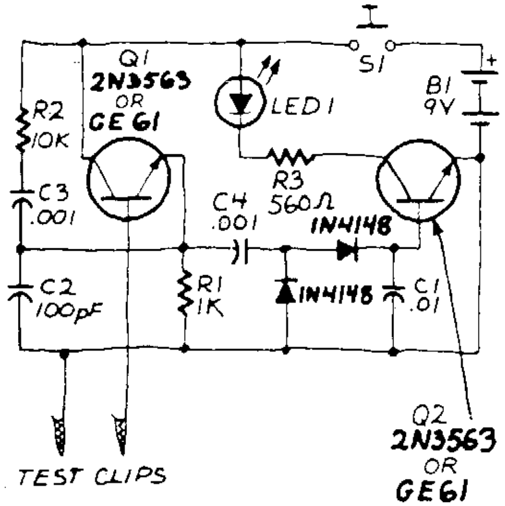
QST

Fig. 19-17

Circuit Notes

A tapped-coil Colpitts oscillator is used at Q1 to provide four tuning ranges from 1.7 to 3.1 MHz, 3.0 to 5.6 MHz, 5.0 to 12 MHz and 11.5 to 31 MHz. A Zener diode (D2) is used at Q1 to lower the operating voltage of the oscillator. A small value capacitor is used at C5 to ensure light coupling to the tuned circuit. Q2 is a source-follower buffer stage. It helps to isolate the oscillator from the generator-output load. The source of Q2 is broadly tuned by means of RFC1. Energy from Q2 is routed to a fed-back, broadband class-A amplifier. A 2 dB attenuator is used at the output of T1 to provide a 50 ohm termination for Q3 and to set the generator-output impedance at 50 ohms. C16, C17 and RFC2 form a brute-force RF decoupling network to keep the generator energy from radiating outside the box on the 12 V supply.

CRYSTAL TESTER



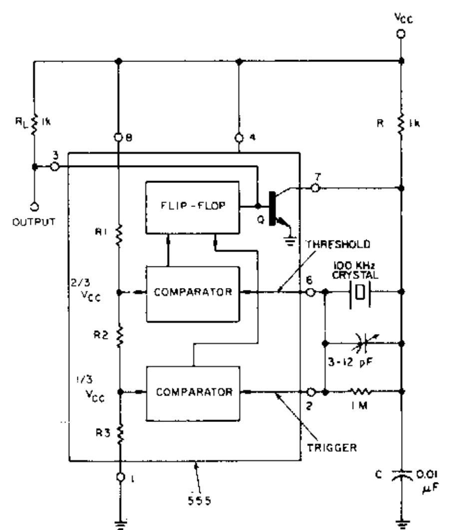
RADIO-ELECTRONICS

Fig. 19-18

Circuit Notes

Transistor Q1, a 2N3563, and its associated components form an oscillator circuit that will oscillate if, and only if, a good crystal is connected to the test clips. The output from the oscillator is then rectified by the two 1N4148 diodes and filtered by C1, a .01 μF capacitor. The positive voltage developed across the capacitor is applied to the base of Q2, another 2N3563, causing it to conduct. When that happens, current flows through LED1, causing it to glow. Since only a good crystal will oscillate, a glowing LED indicates that the crystal is indeed OK. The circuit is powered by a standard nine-volt transistor-radio battery and the SPST pushbutton power-switch is included to prolong battery life.

CRYSTAL-STABILIZED IC TIMER CAN PROVIDE SUBHARMONIC FREQUENCIES



ELECTRONIC DESIGN

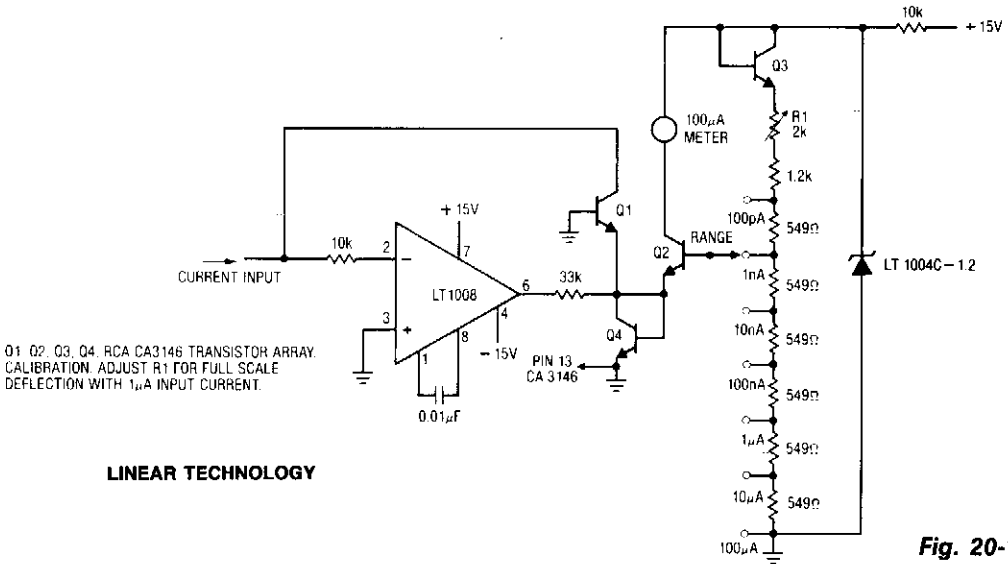
Fig. 19-19

The trimmer across the crystal can finely tune the circuit's oscillating frequency.

20
Current Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AMMETER WITH SIX DECADE RANGE



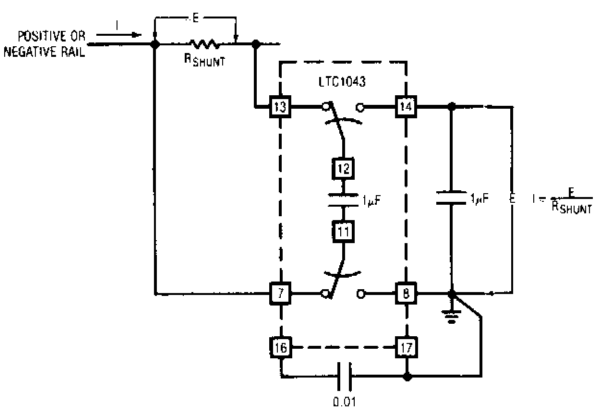
LINEAR TECHNOLOGY

Fig. 20-1

Circuit Notes

The Ammeter measures currents from 100 pA to 100 μA without the use of expensive high value resistors. Accuracy at 100 μA is limited by the offset voltage between Q1 and Q2 and, at 100 pA, by the inverting bias current of the LT1008.

CURRENT SENSING IN SUPPLY RAILS



Circuit Notes

The LTC1043 can sense current through a shunt in either of its supply rails. This capability has wide application in battery and solar-powered systems. If the ground-referred voltage output is unloaded by an amplifier, the shunt can operate with very little voltage drop across it, minimizing losses.

LINEAR TECHNOLOGY CORPORATION *Fig. 20-2*

PICO AMMETER

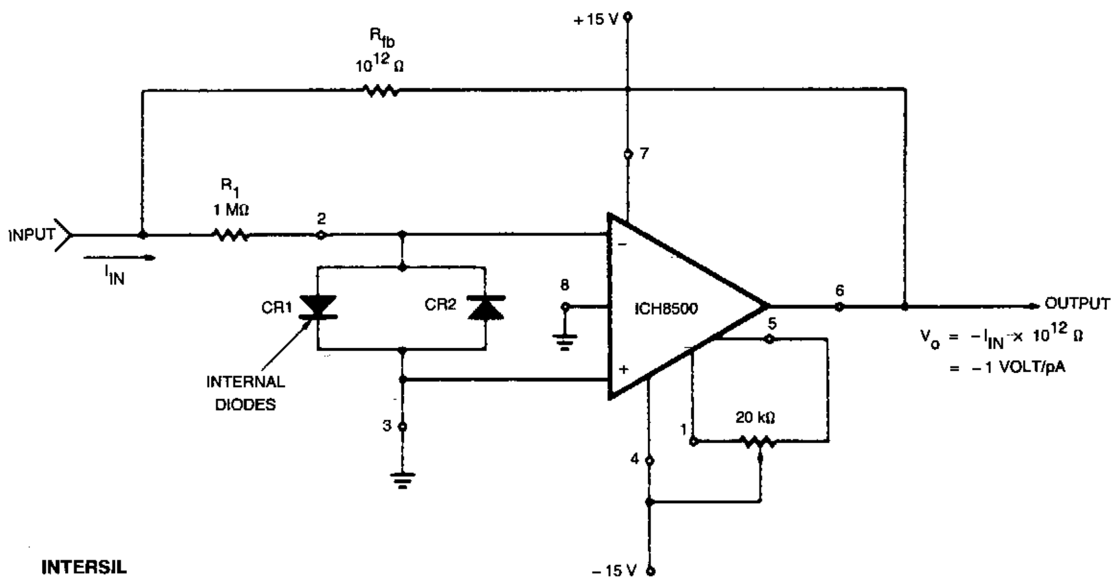
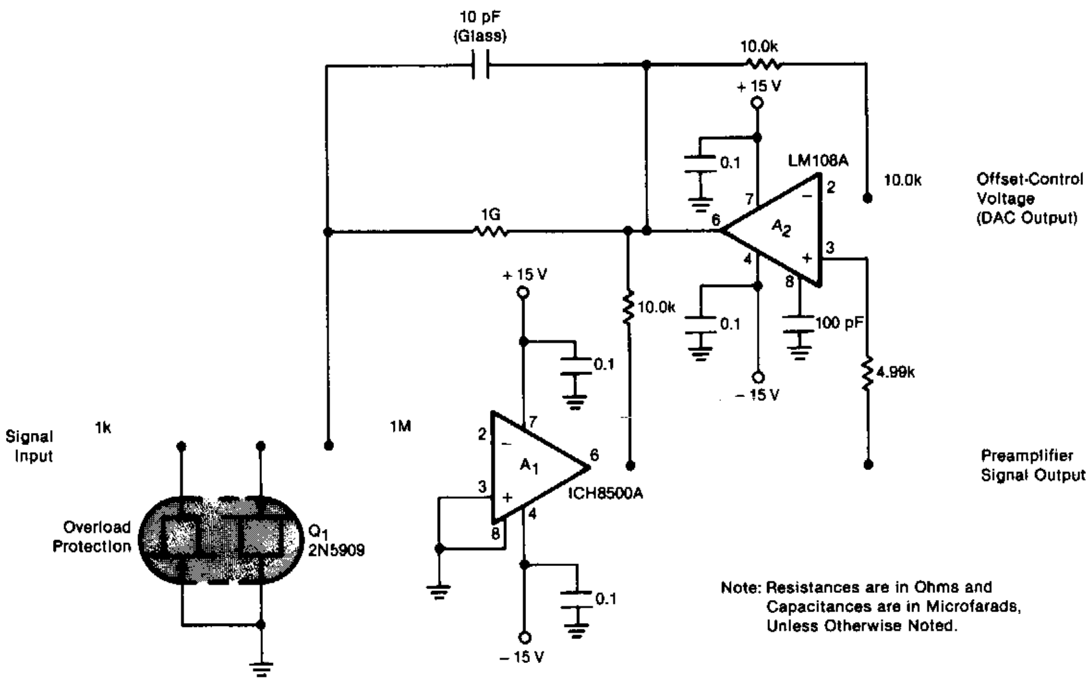


Fig. 20-3

Circuit Notes

Care must be taken to eliminate any stray currents from flowing into the current summing node. This can be accomplished by forcing all points surrounding the input to the same potential as the input. In this case the potential of the input is at virtual ground, or OV. Therefore, the case of the device is grounded to intercept any stray leakage currents that may otherwise exist between the ±15 V input terminals and the inverting input summing junctions. Feedback capacitance should be kept to a minimum in order to maximize the response time of the circuit to step function input currents. The time constant of the circuit is approximately the produce of the feedback capacitance C_{fb} times the feedback resistor R_{fb}. For instance, the time constant of the circuit is 1 sec if C_{fb} = 1pF. Thus, it takes approximately 5 sec (5 time constants) for the circuit to stabilize to within 1% of its final output voltage after a step function of input current has been applied. C_{fb} of less than 0.2 to 0.3 pF can be achieved with proper circuit layout.

ELECTROMETER AMPLIFIER WITH OVERLOAD PROTECTION



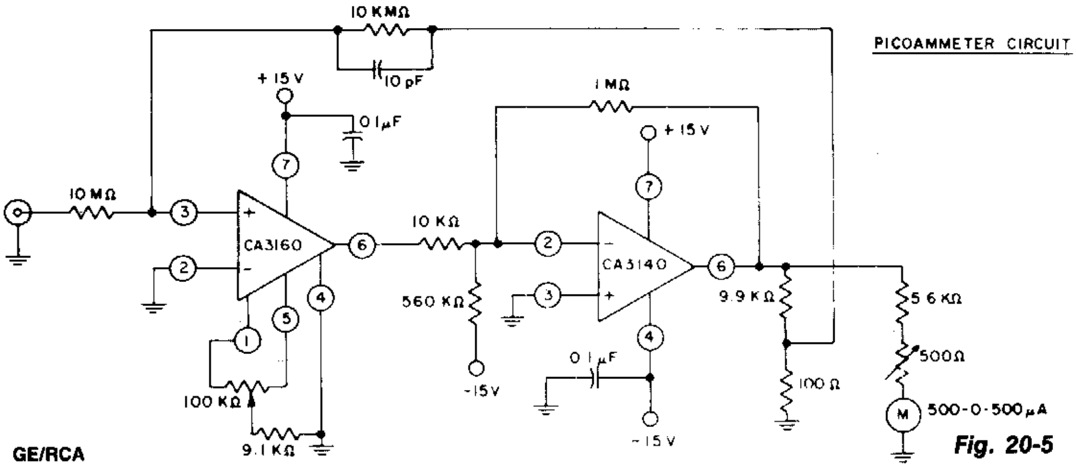
NASA TECH BRIEFS

Fig. 20-4

Circuit Notes

The preamplifier is protected from excessive input signals of either polarity by the 2N5909 junction field-effect transistor. A nulling circuit makes it possible to set the preamplifier output voltage to zero at a fixed low level (up to $\pm 10^{-8}$A) of the input current. (This level is called the standing current and corresponds to the zero-signal level of the instrumentation.) The opposing (offset) current is generated in the 10^9 feedback resistor to buck the standing current. Different current ranges are reached by feeding the preamplifier output to low and high gain amplifier chains. To reduce noise, each chain includes a 1.5 Hz corner active filter.

GUARDED INPUT PICOAMMETER CIRCUIT



GE/RCA

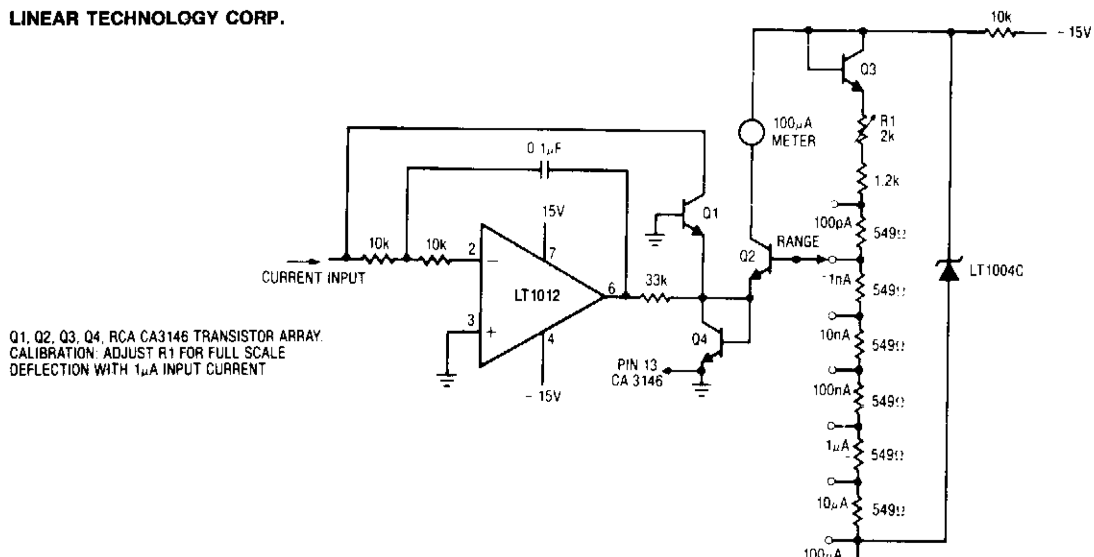
Fig. 20-5

Circuit Notes

The circuit utilizes CA3160 and CA3140 BiMOS op amps to provide a full-scale meter deflection of ±3 pA. The CA3140 serves as an X100 gain stage to provide the required plus and minus output swing for the meter and feedback network. Terminals 2 and 4 of the CA3160 are at ground potential, thus its input is operated in the "guarded mode."

AMMETER WITH SIX DECADE RANGE

LINEAR TECHNOLOGY CORP.

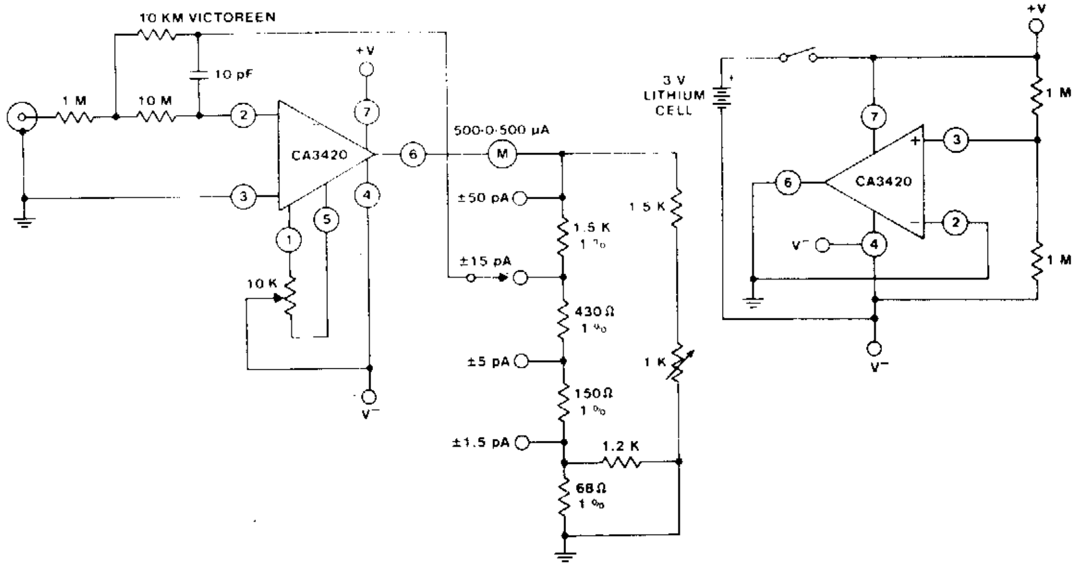


Circuit Notes

Fig. 20-6

The Ammeter measures currents from 100 pA to 100 μA without the use of expensive high value resistors. Accuracy at 100 μA is limited by the offset voltage between Q1 and Q2 and, at 100 pA, by the inverting bias current of the LT1008.

PICOAMMETER CIRCUIT



GENERAL ELECTRIC /RCA *Fig. 20-7*

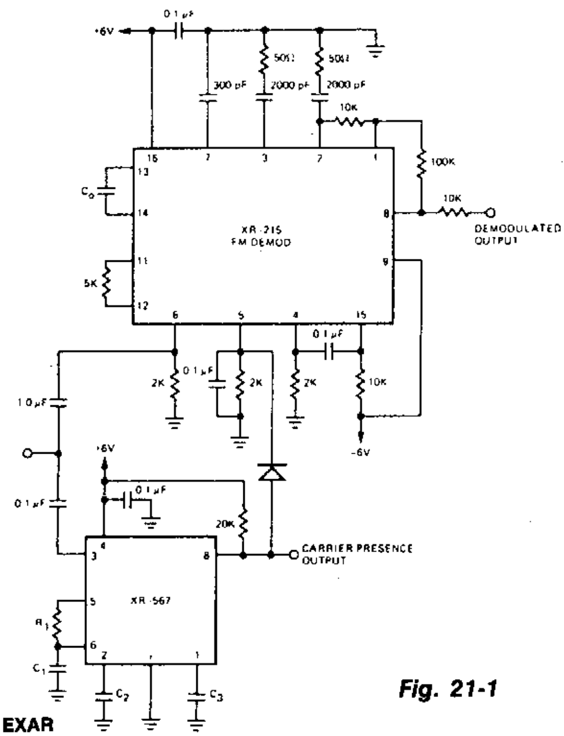
Circuit Notes

The circuit uses the exceptionally low input current (0.1pA) of the CA3420 BiMOS op amp. With only a single 10 megohm resistor, the circuit covers the range from ±50 pA maximum to a full-scale sensitivity of ±1.5 pA. Using an additional CA3420, a low-resistance center tap is obtained from a single 3-volt lithium battery.

21
Demodulators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

NARROW BAND FM DEMODULATOR WITH CARRIER DETECT



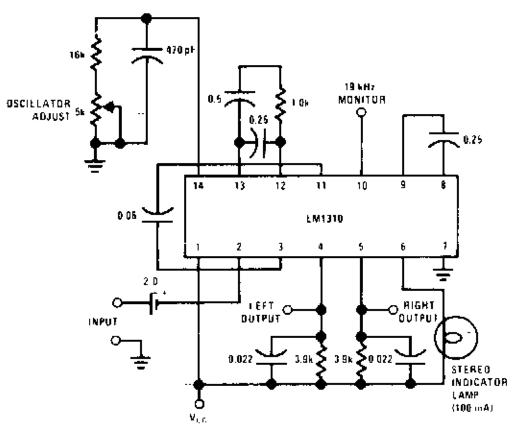
EXAR

Fig. 21-1

Circuit Notes

For FM demodulation applications where the bandwidth is less than 10% of the carrier frequency, an XR-567 can be used to detect the presence of the carrier signal. The output of the XR-567 is used to turn off the FM demodulator when no carrier is present, thus acting as a squelch. In the circuit shown, an XR-215 FM demodulator is used because of its wide dynamic range, high signal/noise ratio and low distortion. The XR-567 will detect the presence of a carrier at frequencies up to 500 kHz.

STEREO DEMODULATOR



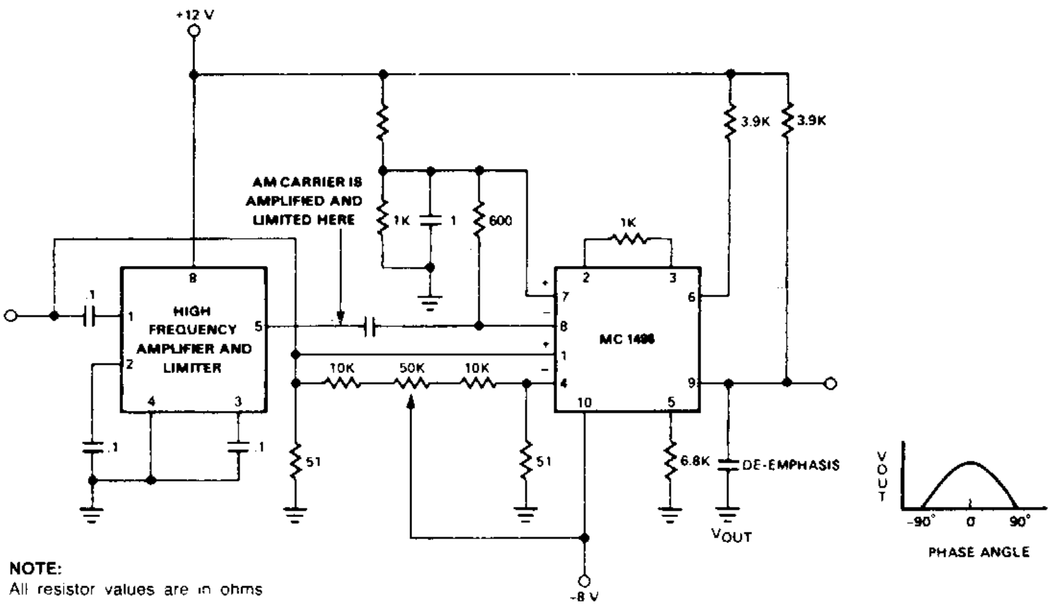
NATIONAL SEMICONDUCTOR CORP.

Fig. 21-2

Circuit Notes

This circuit uses a single IC LM1310 to provide left and right outputs from a composite MPX stereo signal. Oscillator adjust R1 is set for 76 kHz (19 kHz at pin 10). C1 should be a silver mica or NPO ceramic capacitor.

AM DEMODULATOR



NOTE:
All resistor values are in ohms

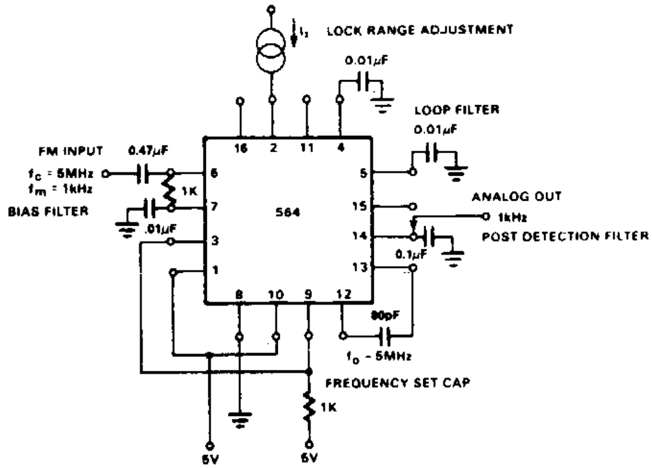
Fig. 21-3

SIGNETICS

Circuit Notes

Amplifying and limiting of the AM carrier is accomplished by the if gain block providing 55 dB of gain or higher with a limiting of 40 μV. The limited carrier is then applied to the detector at the carrier ports to provide the desired switching function. The signal is then demodulated by the synchronous AM demodulator (1496) where the carrier frequency is attentuated due to the balanced nature of the device. Care must be taken not to overdrive the signal input so that distortion does not appear in the recorded audio. Maximum conversion gain is reached when the carrier signals are in phase as indicated by the phase-gain relationship. Output filtering is also necessary to remove high frequency sum components of the carrier from the audio signal.

FM DEMODULATOR



A

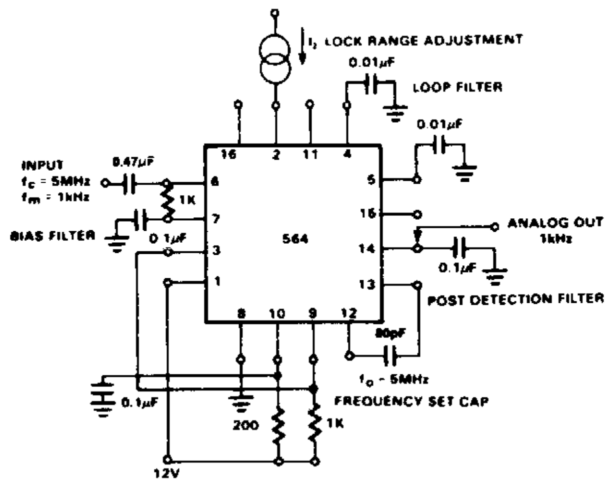


Fig. 21-4

SIGNETICS

B

Circuit Notes

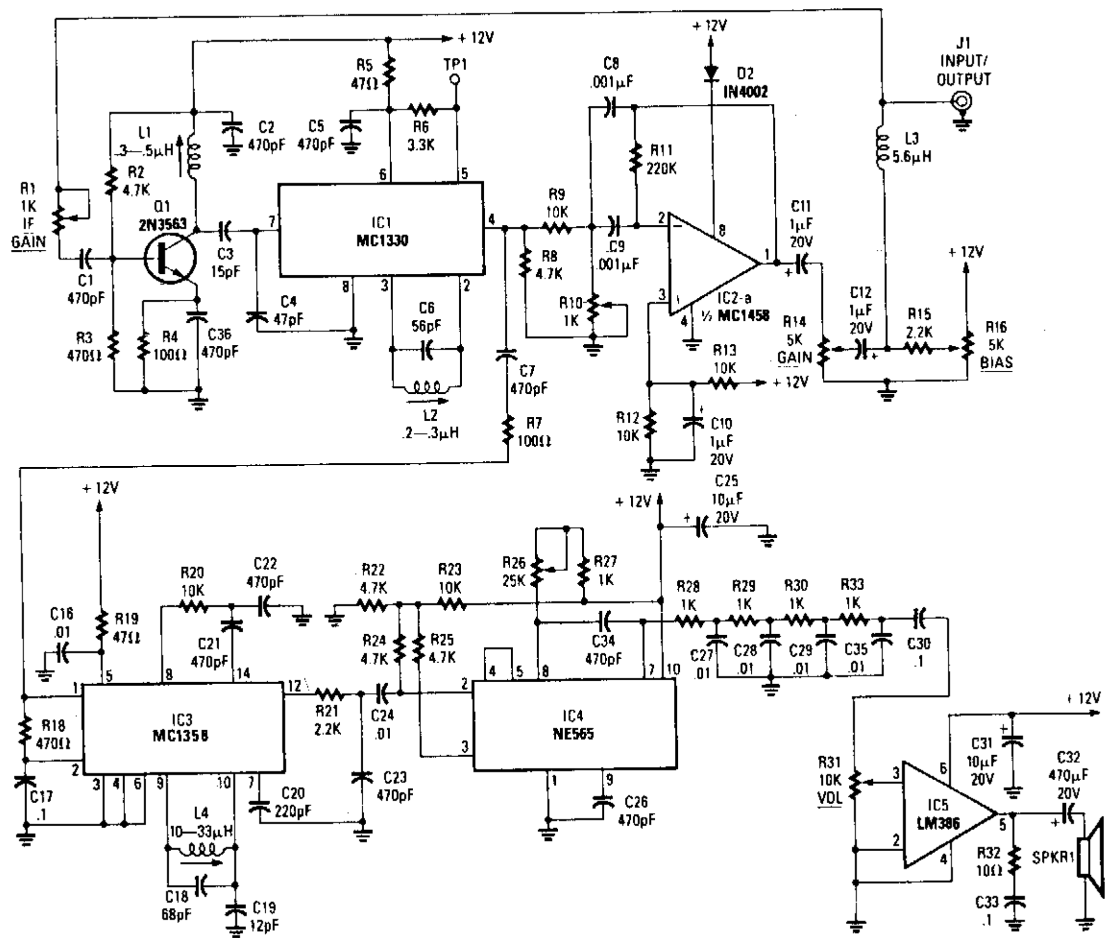
The NE564 is used as an FM demodulator. The connections for operation at 5 V and 12 V are shown in Figures 21-4A and 21-4B. The input signal is ac coupled with the output signal being extracted at Pin 14. Loop filtering is provided by the capacitors at Pins 4 and 5 with additional filtering being provided by the capacitor at Pin 14. Since the conversion gain of the VCO is not very high, to obtain sufficient demodulated output signal the frequency deviation in the input signal should be 1% or higher.

22
Descramblers and Decoders

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Sine Wave Descrambler
Outband Descrambler
Gated Pulse Descrambler
SCA Decoder
Dual Time Constant Tone-Decoder
Stereo TV Decoder
Time Division Multiplex (TDM) Stereo Decoder
Frequency Division Multiplex (FDM) Stereo Decoder
SCA (Background Music) Decoder

SINE WAVE DESCRAMBLER



—A COMPLETE SINEWAVE DESCRAMBLER. Easy to build, and relatively easy to align, this circuit completely removes the 15.75-kHz scrambling sinewave.

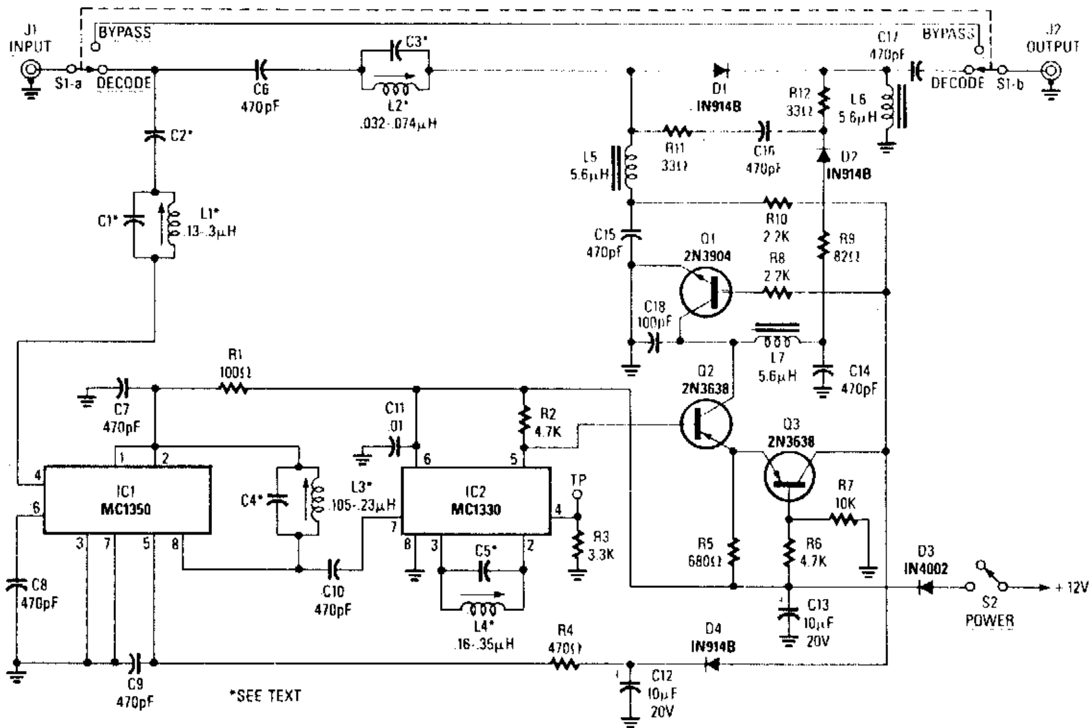
RADIO-ELECTRONICS

Fig. 22-1

Circuit Notes

This decoder features a sine wave recovery channel and uses a PIN diode attenuator driven by the sine wave recovery system to cancel out the sine wave sync suppression signal. Kit available from North Country Radio, P.O. Box 53, Wykagyl Station, New York 10804.

OUTBAND DESCRAMBLER



—FOR THE OUTBAND DECODER shown here to work, the cable company must provide at least a 1 millivolt signal. Values for C1–C5 and L1–L4 are found in Table 1.

TABLE 1—CAPACITOR AND COIL VALUES

	50 MHz	90–114 MHz
C1	5 pF	5 pF
C2	47 pF	12 pF
C3	200 pF	82 pF
C4	56 pF	12 pF
C5	56 pF	10 pF
L1	0.2 μH	0.2 μH
L2	0.05 μH	0.03 μH
L3	0.175 μH	0.2 μH
L4	0.175 μH	0.24 μH

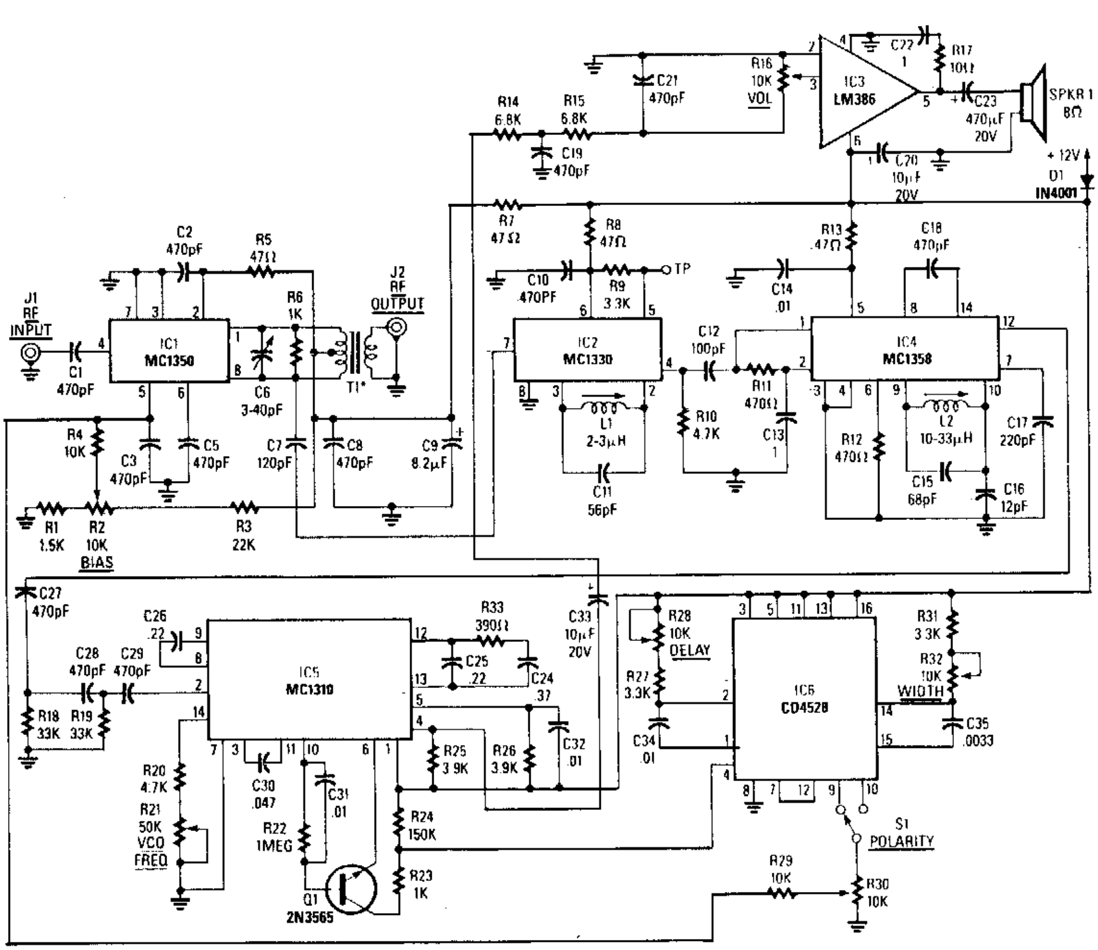
RADIO-ELECTRONICS

Fig. 22-2

Circuit Notes

This circuit consists of an amplifier for the synch channel and a video detector which controls an attenuator so that the gain of the systems is increased during synch intervals. Kit available from North Country Radio, P.O. Box 53, Wykagyl Station, New York 10804.

GATED PULSE DESCRAMBLER



—DESCRAMBLE GATED-PULSE SIGNALS using this easy-to-build circuit. Information for winding transformer T1 and coil L1 can be found in the text.

RADIO-ELECTRONICS

Fig. 22-3

Circuit Notes

This circuit consists of an amplifier and video detector with a second subcarrier detector for synch recovery purposes. A pulse-former circuit modulates the gain of the main channel increasing it during synch intervals. Provision for subcarrier audio descrambling is also provided. Kit available from North Country Radio, P.O. Box 53, Wykagyl Station, New York 10804.

SCA DECODER

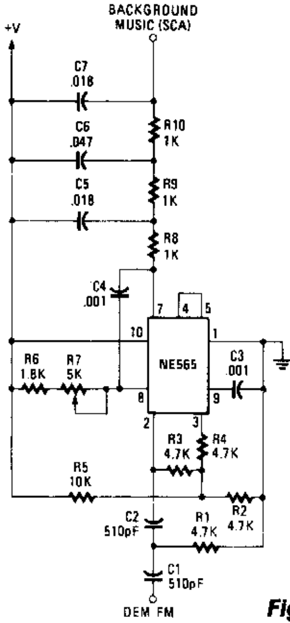


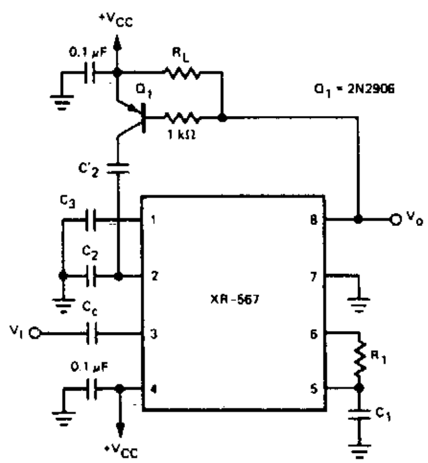
Fig. 22-4

RADIO-ELECTRONICS

Circuit Notes

The circuit uses a Signetics NE565 PLL (Phase-Locked Loop) as a detector to recover the SCA signal. The input to the SCA decoder circuit is connected to an FM receiver at a point between the FM discriminator and the de-emphasis filter network. The PLL, IC1, is tuned to 67 kHz by R7, a 5 K potentiometer. Tuning need not be exact since the circuit will seek and lock onto the subcarrier. The demodulated signal from the FM receiver is fed to the input of the 565 through a high-pass filter consisting of two 510 pF capacitors (C1 and C2) and a 4.7 K resistor (R1). Its purpose is to serve as a coupling network and to attenuate some of the main channel spill. The demodulated SCA signal at pin 7 passes through a three-stage de-emphasis network as shown. The resulting signal is around 50 mV, with the response extending to around 7 kHz.

DUAL TIME CONSTANT TONE DECODER



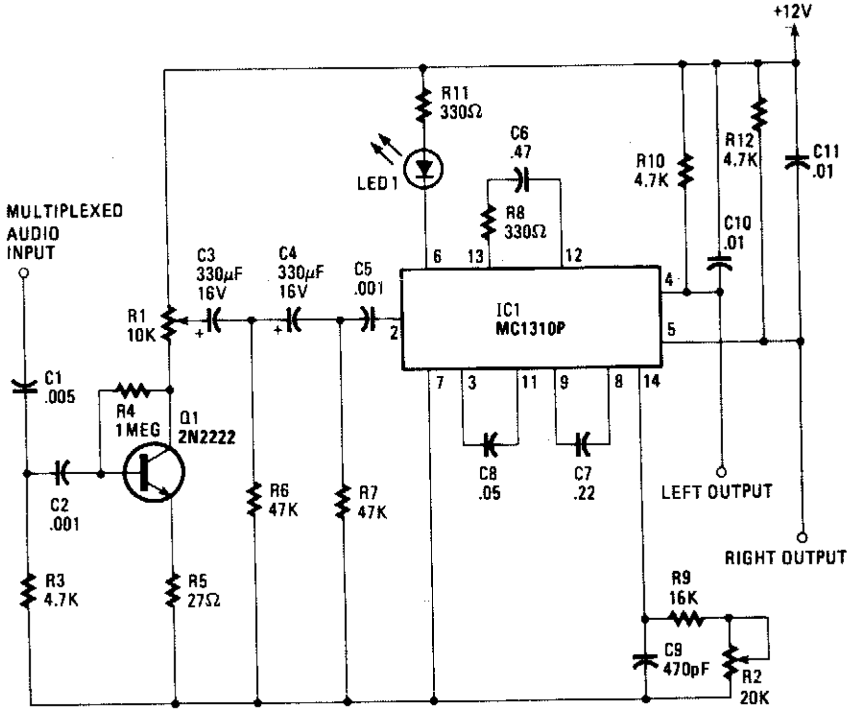
EXAR

Fig. 22-5

Circuit Notes

For some applications it is important to have a tone decoder with narrow bandwidth and fast response time. This can be accomplished by the dual time constant tone decoder circuit shown. The circuit has two low-pass loop filter capacitors, C_2 and C'_2. With no input signal present, the output at pin 8 is high, transistor Q_1 is off, and C'_2 is switched out of the circuit. Thus, the loop low-pass filter is comprised of C_2, which can be kept as small as possible for minimum response time. When an in-band signal is detected, the output at pin 8 will go low, Q_1 will turn on, and capacitor C'_2 will be switched in parallel with capacitor C_2. The low-pass filter capacitance will then be $C_2 + C'_2$. The value of C'_2 can be quite large in order to achieve narrow bandwidth. During the time that no input signal is being received, the bandwidth is determined by capacitor C_2.

STEREO TV DECODER



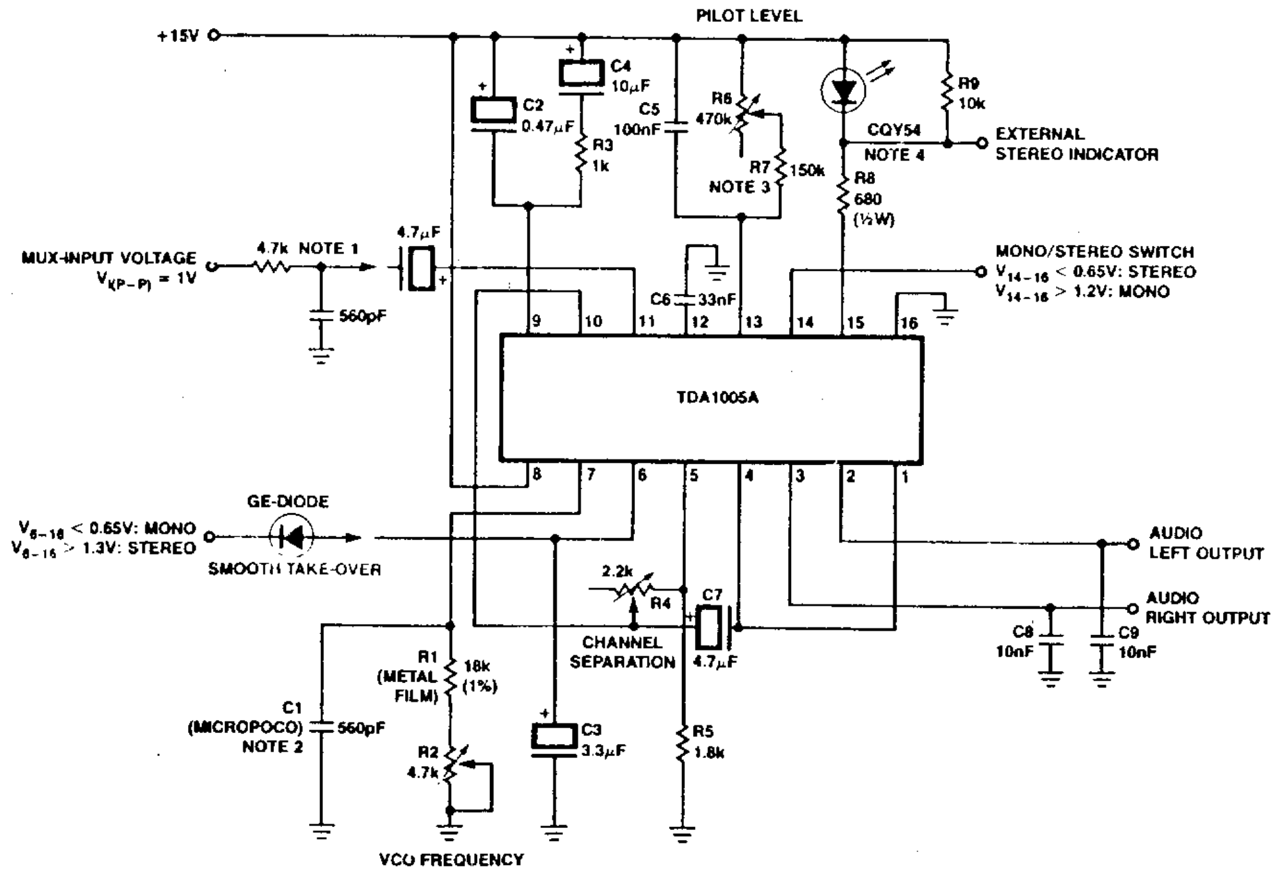
RADIO-ELECTRONICS

Fig. 22-6

Circuit Notes

The composite input signal is preamplified by transistor Q1 and is then coupled to the high-pass filter composed of C3, C4, R6, and R7. The filtered audio is then passed to IC1, an MC131OP "Coilless Stereo Demodulator." That IC is normally used to demodulate broadcast-band FM signals, but by changing the frequency of its on-board VCO (Voltage Controlled Oscillator) slightly (from 19 kHz to 15.734 kHz), we can use that IC to detect stereo-TV signals. A block diagram of the MC131OP is shown in Fig. 22-5. Notice that the components connected to pin 14 control the VCO's frequency, hence the pilot-detect and carrier frequencies. For use in an FM receiver, the VCO would run at four times the 19 kHz pilot frequency (76 kHz), but for our application, it will run at four times the 15.734 kHz pilot frequency of stereo TV, or 62.936 kHz. The MC131OP divides the master VCO signal by two in order to supply the 31.468 kHz carrier that is used to detect the L − R audio signal. The L − R signal undergoes normal FM detection, and at that point we've got two audio signals: L + R and L − R. The decoder block in the IC performs the addition and subtraction to produce the separate left and right signals. R10 and C10 form a de-emphasis network that compensates for the 75 μs pre-emphasis that the left channel underwent; R12 and C11 perform the same function for the right channel.

TIME DIVISION MULTIPLEX (TDM) STEREO DECODER

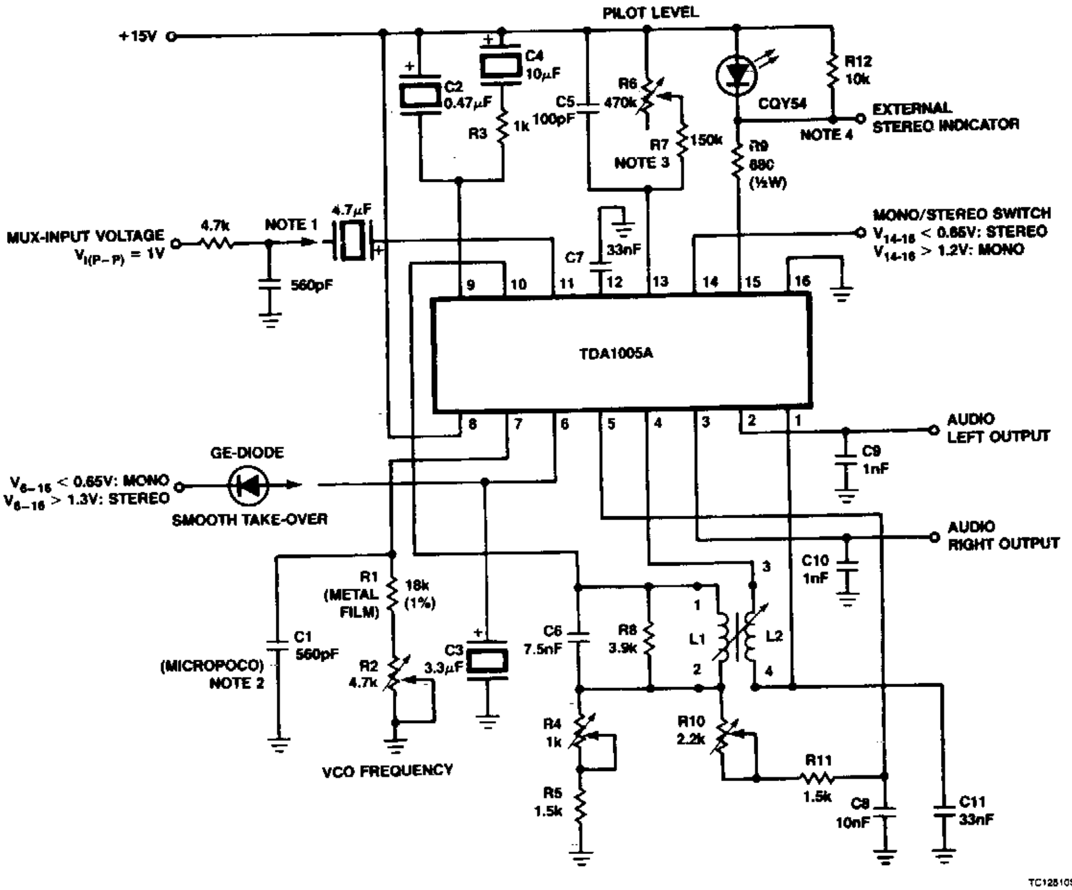


NOTES:
1. For other input structures see Figures 7 to 11, shown here with RC-filter (Figure 8).
2. Micropoco capacitor has a temperature coefficient of $125.10^{-6} \pm 60.10^{-6}{}^{\circ}C^{-1}$.
3. In simplified circuits a fixed resistor (e.g. 620k) can be used for a guaranteed switching level of $\leqslant$ 16mV.
4. Either the LED circuit or an external stereo indicator can be used.

SIGNETICS

Fig. 22-7

FREQUENCY DIVISION MULTIPLEX (FDM) STEREO DECODER



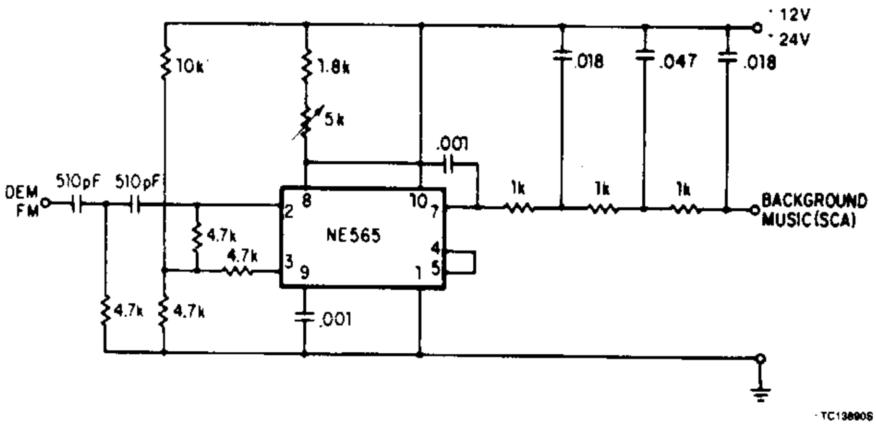
Coil data:
L_1 L_2 = 2.6mH
Q_{1-2} = 35; Q_{MIN} = 30
N_{1-2} = 357½ turns;
N_{3-4} = 297½ turns: scrambled wound with wire diameter 0.09mm, $\frac{E_{3-4}}{E_{1-2}} \times 100\% = 82\%$

NOTES:
1. For other input structures see Figures 6 to 11; shown here with RC-filter (Figure 8).
2. The micropoco capacitor has a temperature coefficient of $125.10^{-6} \pm 60.10^{-6}$ k^{-1}.
3. In simplified circuits a fixed resistor (e.g. 620k) can be used for a guaranteed switching level of ≤ 16mV.
4. Either the LED circuit or an external stereo indicator can be used.

SIGNETICS

Fig. 22-8

SCA (Background Music) DECODER



SIGNETICS

Fig. 22-9

Circuit Notes

A resistive voltage divider is used to establish a bias voltage for the input (Pins 2 and 3). The demodulated (multiplex) FM signal is fed to the input through a two-stage high-pass filter, both to effect capacitive coupling and to attenuate the strong signal of the regular channel. A total signal amplitude, between 80 mV and 300 mV, is required at the input. Its source should have an impedance of less than 10,000 ohm. The Phase-Locked Loop is tuned to 67 kHz with a 5000 ohm potentiometer, only approximate tuning is required since the loop will seek the signal. The demodulated output (Pin 7) passes through a three-stage low-pass filter to provide de-emphasis and attenuate the high-frequency noise which often accompanies SCA transmission. Note that no capacitor is provided directly at Pin 7; thus, the circuit is operating as a first-order loop. The demodulated output signal is in the order of 50 mV and the frequency response extends to 7 kHz.

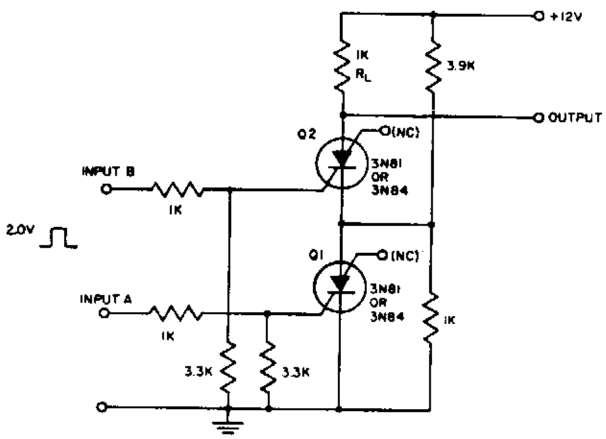
23

Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Pulse Sequence Detector
Voltage Level Detector
Zero-Crossing Detector
Peak Detector
Level Detector
High Frequency Peak Detector
Tachometer, Single Pulse Generator, Power Loss
  Detector, Peak Detector
Phase Detector with 10-Bit Accuracy
Frequency Limit Detector
Pulse Coincidence Detector

PULSE SEQUENCE DETECTOR



GENERAL ELECTRIC

Fig. 23-1

Circuit Notes

The resistor divider connected between Q1 and Q2 supplies I_H to Q1 after input A triggers it. It also prevents input B from triggering Q2 until Q1 conducts. Consequently, the first B input pulse after input A is applied will supply current to R_L.

VOLTAGE LEVEL DETECTOR

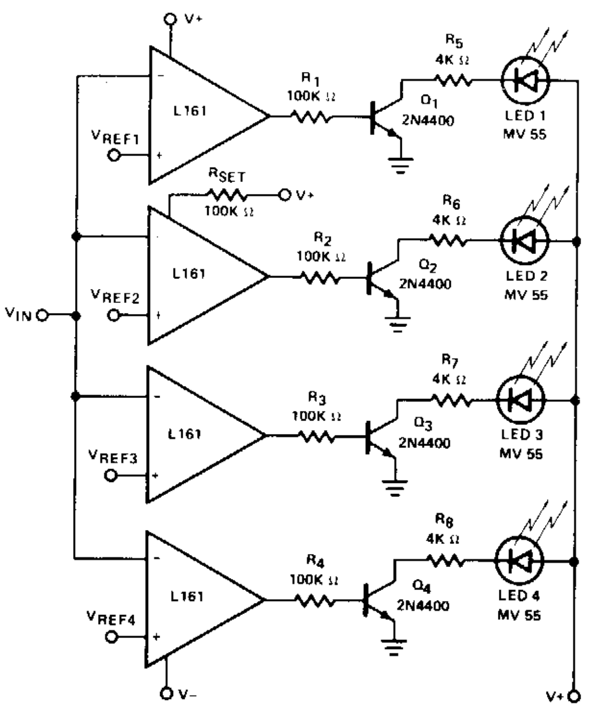
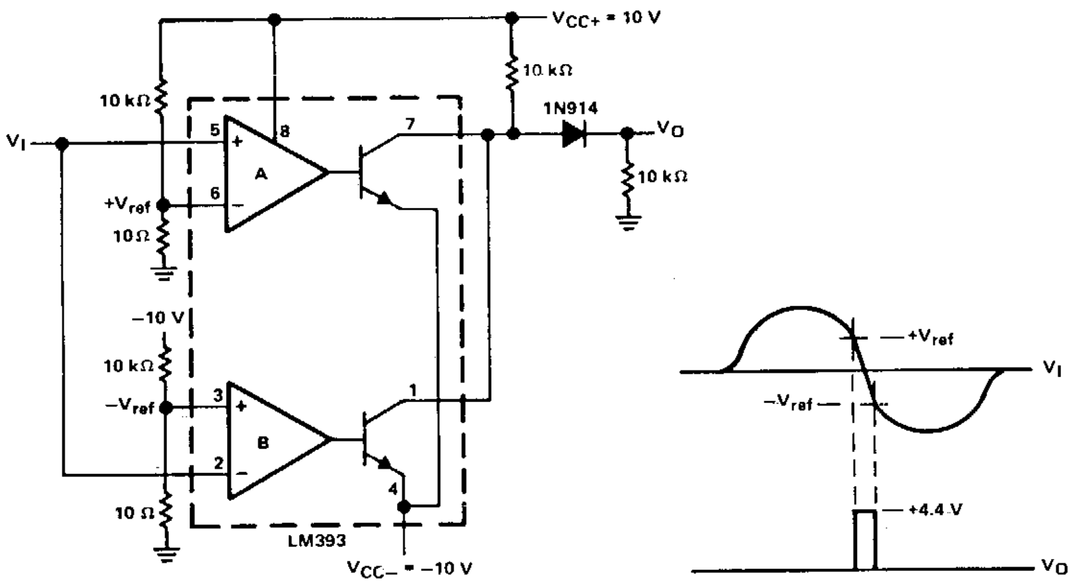


Fig. 23-2

SILICONIX

ZERO-CROSSING DETECTOR



TEXAS INSTRUMENTS

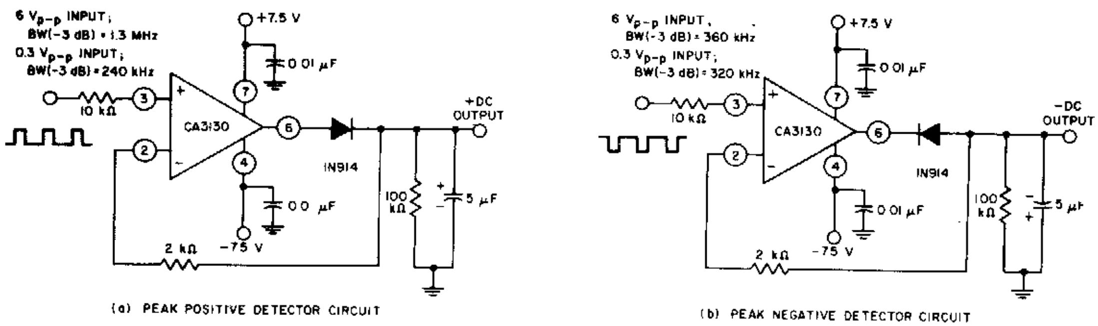
Fig. 23-3

Circuit Notes

This zero-crossing detector uses a dual LM393 comparator, and easily controls hysteresis by the reference levels which are set on the comparator inputs. The circuit illustrated is powered by ±10-V power supplies. The input signal can be an ac signal level up to +8 V. The output will be a positive going pulse of about 4.4 V at the zero-crossover point. These parameters are compatible with TTL logic levels.

The input signal is simultaneously applied to the non-inverting input of comparator A and the inverting input of comparator B. The inverting input of comparator A has a +10 mV reference with respect to ground, while the non-inverting input of comparator B has a −10 mV reference with respect to ground. As the input signal swings positive (greater than +10 mV), the output of comparator "A" will be low while comparator "B" will have a high output. When the input signal swings negative (less than −10 mV), the reverse is true. The result of the combined outputs will be low in either case. On the other hand, when the input signal is between the threshold points (±10 mV around zero crossover), the output of both comparators will be high. If more hysteresis is needed, the ±10 mV window may be made wider by increasing the reference voltages.

PEAK DETECTOR



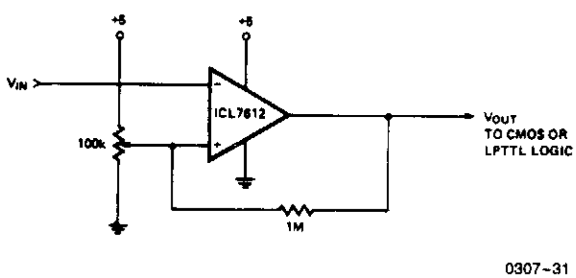
GENERAL ELECTRIC/RCA

Fig. 23-4

Circuit Notes

Circuits are easily implemented using the CA3130 BiMOS op amp. For large-signal inputs the bandwidth of the peak-negative circuit is less than that of the peak-positive circuit. The second stage of the CA3130 limits bandwidth in this case.

LEVEL DETECTOR



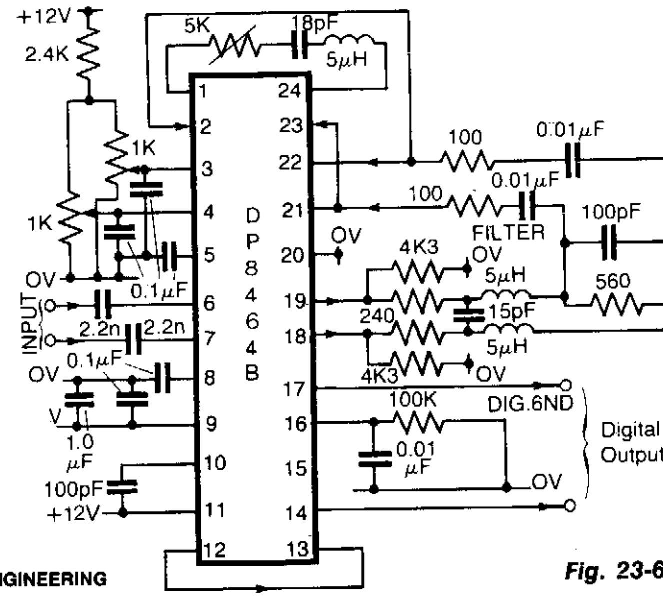
INTERSIL

Fig. 23-5

Circuit Notes

By using the ICL7612 in these applications, the circuits will follow rail to rail inputs.

HIGH FREQUENCY PEAK DETECTOR



ELECTRONIC ENGINEERING

Fig. 23-6

Circuit Notes

National Semiconductor's DP8464B is primarily intended for use in disk systems as a pulse detector. However it can be easily used as a general purpose peak detector for analogue signals up to 5 MHz. The chip can handle signals between 20 and 66 mV peak-to-peak. The circuit includes a filter with constant group delay characteristics to band limit the signal. Typically the −3 dB point for this filter will be at about 1.5 times the highest frequency of interest. This differentiator network between pins 1 and 24 can be as simple as a capacitor, or can be more complex to band limit the differentiator response.

TACHOMETER, SINGLE PULSE GENERATOR, POWER LOSS DETECTOR, PEAK DETECTOR

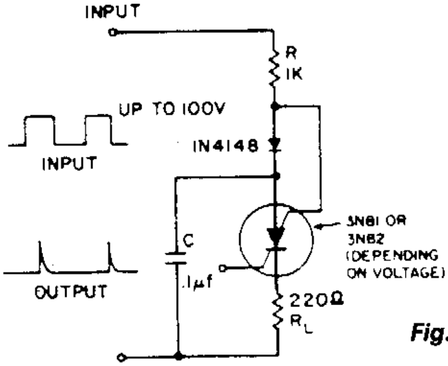
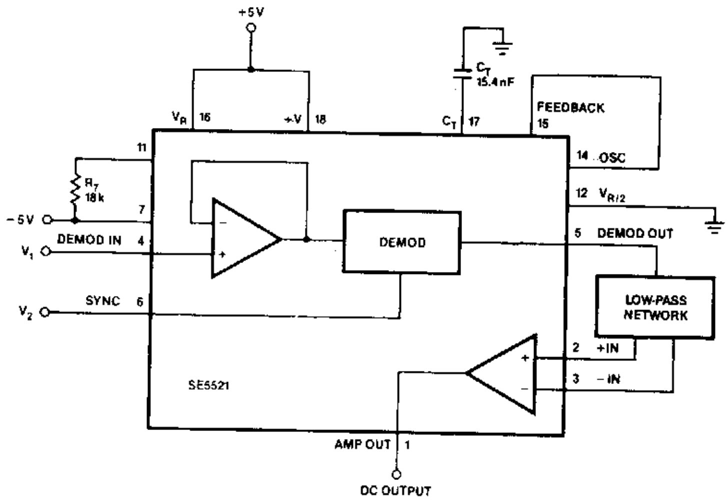


Fig. 23-7

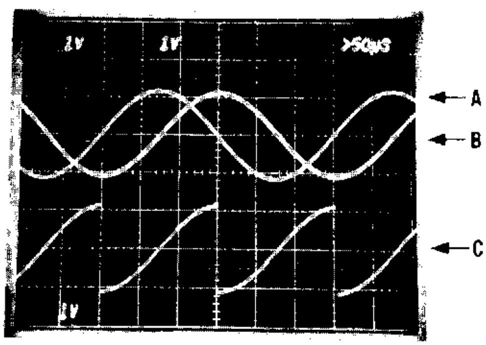
GENERAL ELECTRIC

A positive going input charges C through the IN4148 and R. The diode keeps the scs off. A negative going input supplies anode-gate current triggering on the scs discharging C through R_L.

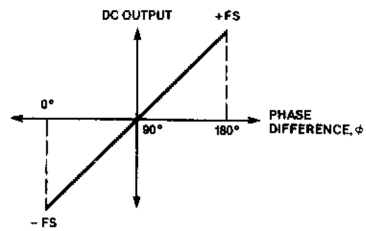
PHASE DETECTOR WITH 10-BIT ACCURACY



a. Phase Detector Measures Phase Difference Between Signals V_1 and V_2 and Provides dc Output at Pin 1



b. When V_1 and V_2 in (a) are at Quadrature (Traces A and B), the DC Component of Demodulator Output (Trace C) is at 0V



c. The dc Output and Phase Vary Linearly

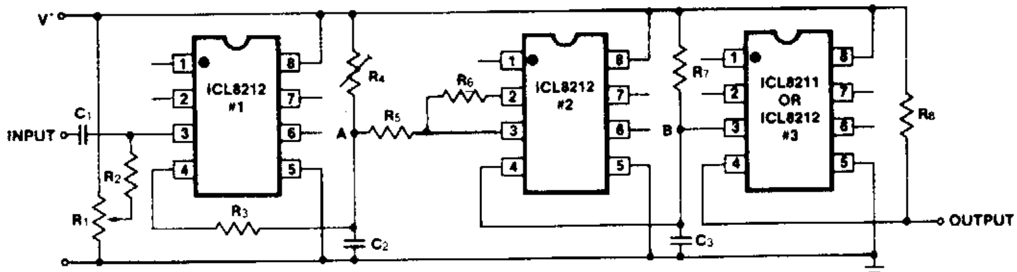
SIGNETICS

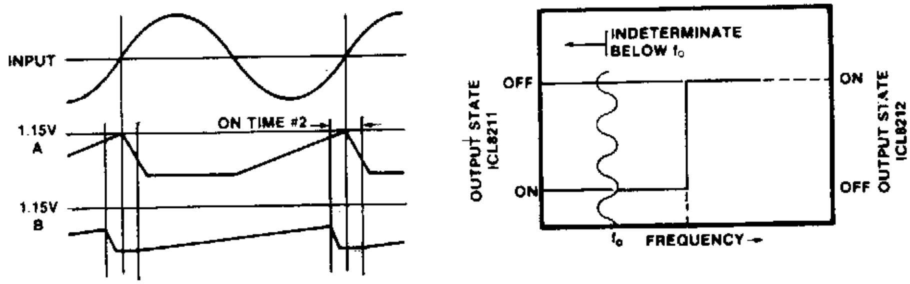
Fig. 23-8

Circuit Notes

Signals of identical frequency are applied to sync input (Pin 6) and to the demodulator input (Pin 4), respectively, the demodulator functions as a phase detector with output dc component being proportional to phase difference between the two inputs. The signals must be referenced to 0 V for dual supply operation or to $V_R/2$ for single supply operation. At ±5-V supplies, the demodulator can easily handle 7-V peak-to-peak signals. The low-pass network configured with the uncommitted amplifier dc output at Pin 1 of the device. The dc output is maximum (+ full-scale) when V_1 and V_2 are 180° out of phase and minimum (− full-scale) when the signals are in phase.

FREQUENCY LIMIT DETECTOR





INTERSIL

Fig. 23-9

Circuit Notes

Simple frequency limit detectors providing a GO/NO-GO output for use with varying amplitude input signals may be conveniently implemented with the ICL8211/8212. In the application shown, the first ICL8212 is used as a zero-crossing detector. The output circuit consisting of R_3, R_4 and C_2 results in a slow output positive ramp. The negative range is much faster than the positive range. R_5 and R_6 provide hysteresis so that under all circumstances the second ICL8212 is turned on for sufficient time to discharge C_3. The time constant of R_7C_3 is much greater than R_4C_2. Depending upon the desired output polarities for low and high input frequencies, either an ICL8211 or an ICL8212 may be used as the output driver.

The circuit is sensitive to supply voltage variations and should be used with a stabilized power supply. At very low frequencies the output will switch at the input frequency.

PULSE COINCIDENCE DETECTOR

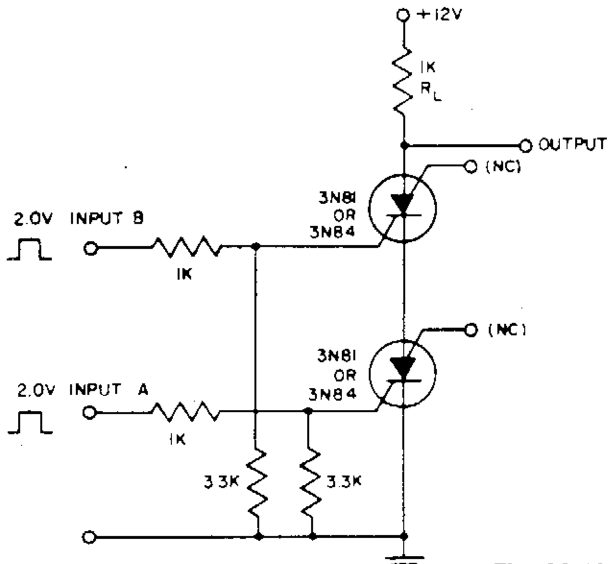


Fig. 23-10

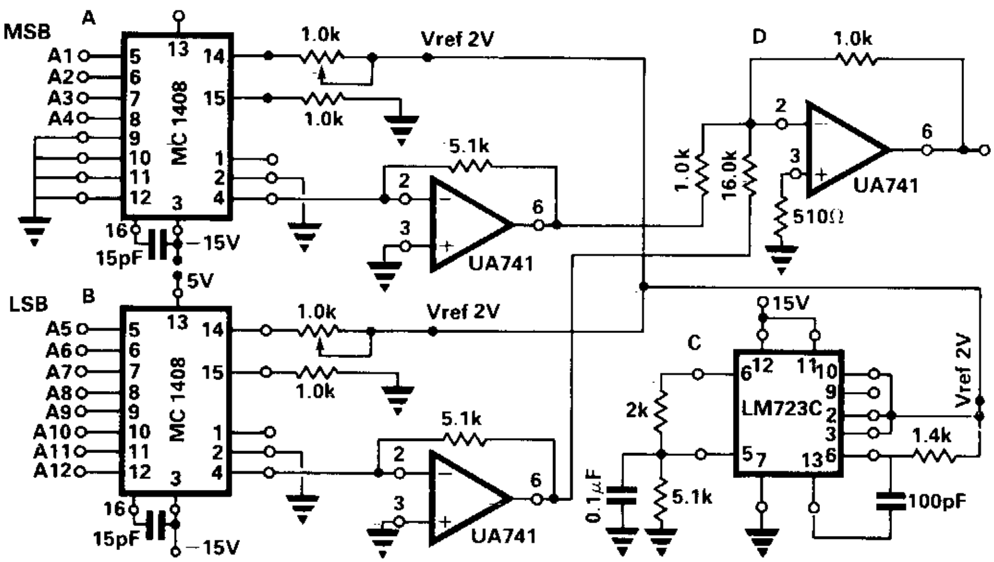
Circuit Notes

Unless inputs A and B (2- to 3-V amplitude) occur simultaneously no voltage exists across R_L. Less than 1 microsecond overlap is sufficient to trigger the scs. Coincidence of negative inputs is detected with gates G_A instead of G_C by using the scs in a complementary SCR configuration.

24
Digital-to-Analog Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TWO 8-BIT DACS MAKE A 12-BIT DAC



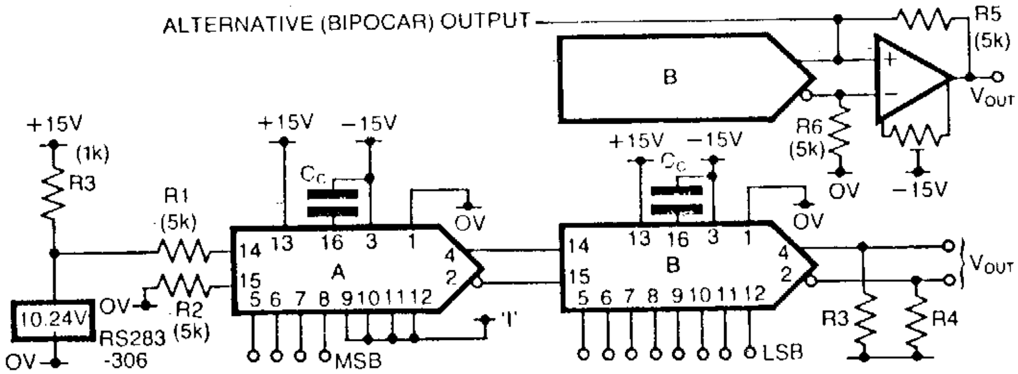
ELECTRONIC ENGINEERING

Fig. 24-1

Circuit Notes

Two MC1408—8-bit D/A converters, A and B in the circuit diagram, are used. The four least-significant bits of A are tied to zero. The four most significant bits of the 12-bit data are connected to the remaining four input pins. The eight least significant bits of the 12-bit data are connected to the eight input pins of B. The four most significant bits of the 12-bit data together have a weight of 16 relative to the remaining eight bits. Hence, the output from B is reduced by a factor of 16 and summed with the output from A using the summing op-amp configuration D. Voltage regulator chip, LM7236, is used to provide an accurate reference voltage, 2 V, for the MC1408. The full-scale voltage of the converter is $\frac{1}{16} \times 9.9609 + 1 \times (9.375)$ or 9.9976 V. The step size of the converter is 2.4 mV.

12-BIT DAC WITH VARIABLE STEP SIZE



ELECTRONIC ENGINEERING

Fig. 24-2

Circuit Notes

The step size of the converter is variable by selection of the high order data bits. The first DAC, A, has a stable reference current supplied via the 10.24 V reference IC and R1. R2 provides bias cancellation. As shown, only the first 4 MSB inputs are used, giving a step size of 225/256 × 2.048/16 = 0.127 mA. This current supplies the reference for DAC B whose step size is then 0.1275/256 = 0.498 μA. Complementary voltage outputs are available for unipolar output and using R3 = R4 = 10 K, V_{out} is ±10.2 V approximately, with a step size (1 LSB) of approximately 5 mV. If desired an op amp can be added to the output to provide a low impedance output with bipolar output symmetrical about ground, if R5 = R6 within 0.05%. Note that offset null is required, and all resistors except R2 and R3 should be 1% high stability types.

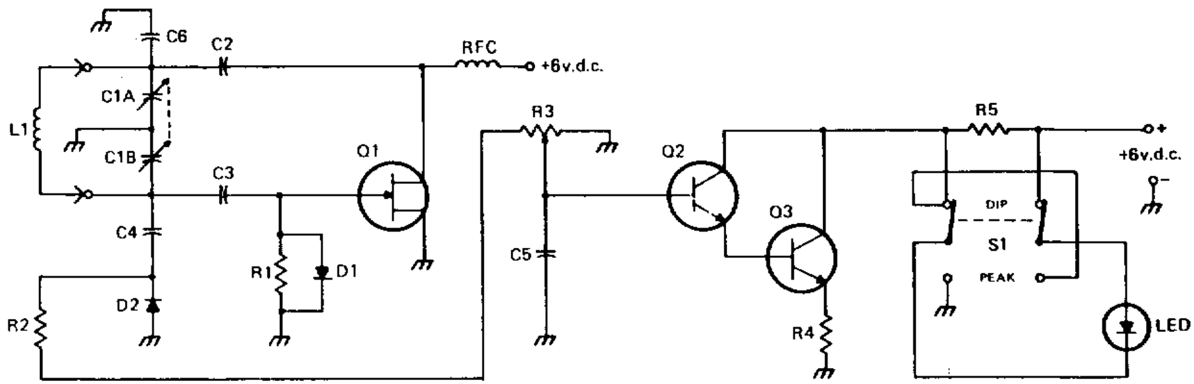
By using lower order address lines than illustrated for DAC A, a smaller step size (and therefore full-scale output) can be obtained. Unused high order bits can be manipulated high or low to change the relative position of the full-scale output.

25
Dip Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Little Dipper

LITTLE DIPPER



CQ

Parts List

L1—See coil data.
C1A,1B—Dual capacitor 100 pF per section (ETCO SV409 or similar)
C2,C3—100 pF mica, mylar, etc., low voltage
C4—10 pF mica, mylar, etc., low voltage
C5—.01 uF ceramic, low voltage
C6—5 pF mica, mylar, etc., low voltage
D1,D2—1N914 silicon diode or similar
R1—100 K ohms ¼ watt
R2—220 K ohms ¼ watt
R3—500 K ohms potentiometer
R4—10 ohms ¼ watt
R5—270 ohms ¼ watt
Q1—MPF 102 FET
Q2—Any general-purpose NPN transistor with a Beta (Hfe) of 40 or so (2N3904 or similar)
Q3—Any general-purpose NPN transistor capable of 20 mA collector current or more, Beta 40 or so (2N3904, 2N2222 or similar)
RFC—1 mH miniature ferrite core choke (value not critical)
LED—Panel mounting LED Radio Shack 276-068 or similar.
SW1—Sub-miniature DPDT slide switch or similar

Miscellaneous—6 volt AC adapter (Radio Shack 273-1454A)
Coaxial DC power jack (RS 274-1565)
Calibrated Dial knob (RS 274-413)
Dual phono jack (RS 274-332)

Fig. 25-1

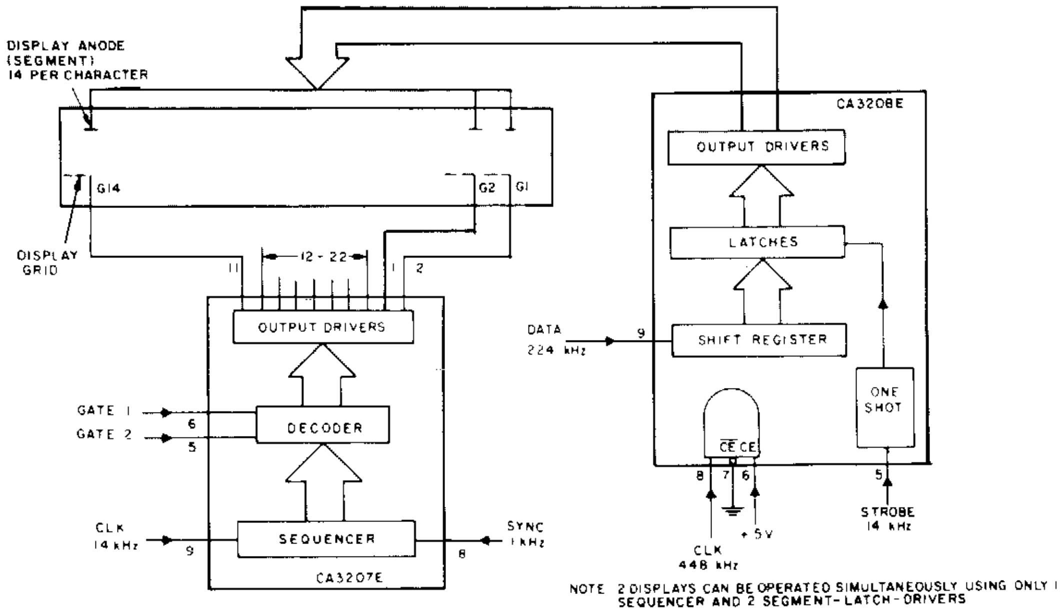
Circuit Notes

The circuit consists of two basic circuits, the oscillator and the detector. The oscillator uses an FET in a Colpitts configuration. The energy circulating in the oscillator tank is coupled through C4 to the detector circuit, where a small diode (D2) rectifies it, feeding a dc voltage to the Darlington pair (Q2, A3) controlled by the sensitivity control (R3). Any small variations in the bias of the amplifier will cause large variations of current through the LED indicator in the DIP mode; however, in the PEAK mode the current produces a corresponding voltage drop through R5 and the action of the LED is reversed. The circuit shown will work practically on any frequency from LF to VHF if the appropriate components are used.

26
Display Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

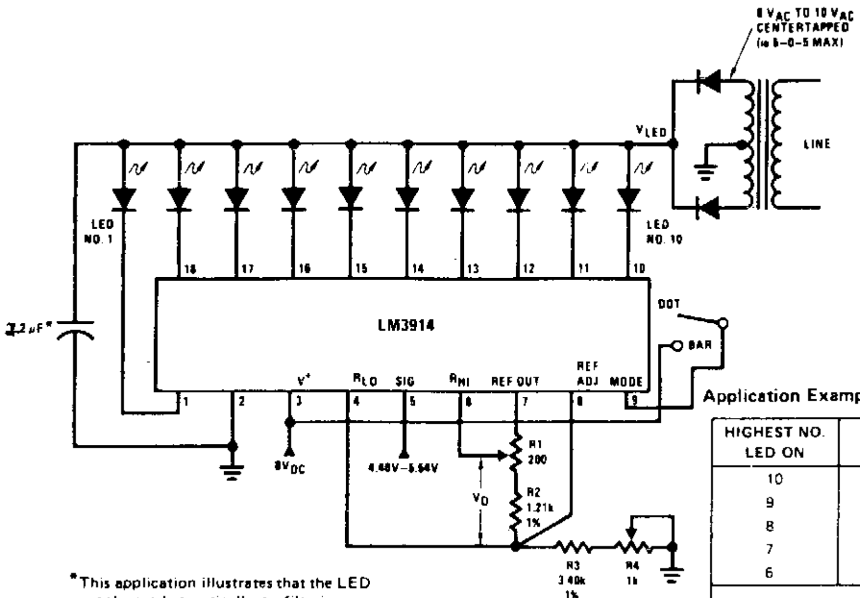
VACUUM FLUORESCENT DISPLAY



GENERAL ELECTRIC/RCA *Fig. 26-1*

This circuit uses the CA3207 sequence driver and CA3208 segment latch-driver in combination to drive display devices of up to 14 segments with up to 14 characters of display. The CA3207 selects the digit or character to be displayed in sequence, CA3208 turns on the required alphanumeric segments.

EXPANDED SCALE METER, DOT OR BAR



*This application illustrates that the LED supply needs practically no filtering

Calibration: With a precision meter between pins 4 and 6 adjust R1 for voltage V_D of 1.20V. Apply 4.94V to pin 5, and adjust R4 until LED No. 5 just lights. The adjustments are non-interacting.

Application Example: Grading 5V Regulators

HIGHEST NO. LED ON	COLOR	VOUT(MIN)
10	Red	5.54
9	Red	5.42
8	Yellow	5.30
7	Green	5.18
6	Green	5.06
	5V	
5	Green	4.94
4	Green	4.82
3	Yellow	4.7
2	Red	4.58
1	Red	4.46

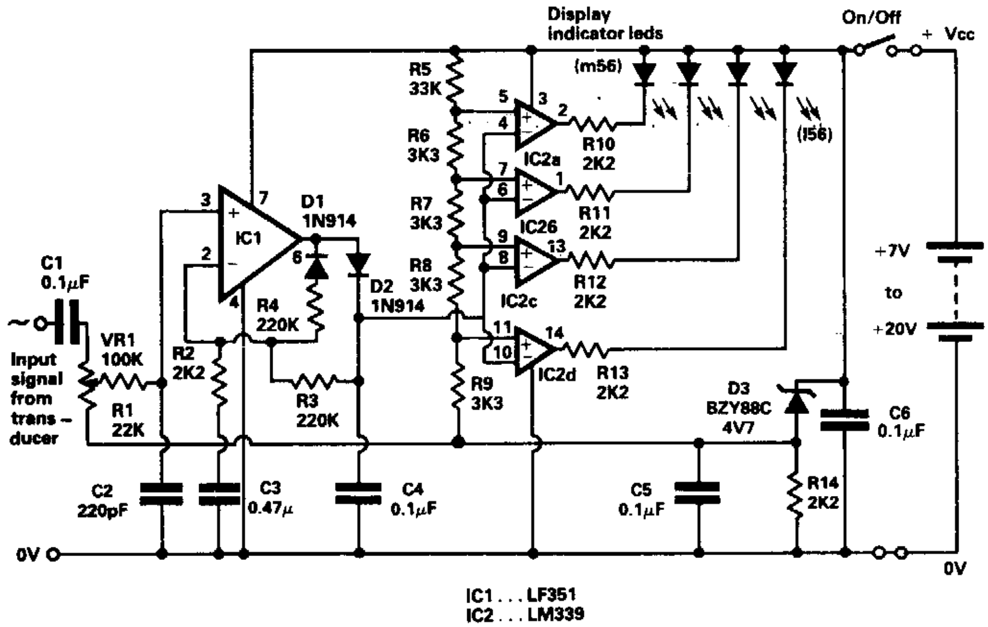
NATIONAL SEMICONDUCTOR CORP.

Fig. 26-2

Circuit Notes

A bar graph driver IC LM314 drives an LED display. The LEDs may be separate or in a combined (integral) bar graph display.

LOW-COST BAR-GRAPH INDICATOR FOR AC SIGNALS



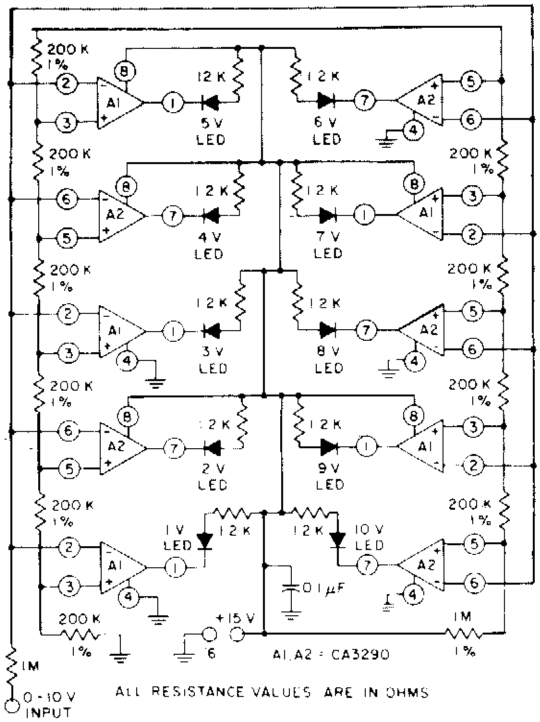
ELECTRONIC ENGINEERING

Fig. 26-3

Circuit Notes

Indicator was designed for displaying the peak level of small ac signals from a variety of transducers including microphones, strain gauges and photodiodes. The circuit responds to input signals contained within the audio frequency spectrum, i.e., 30 Hz to 20 kHz, although a reduced response extends up to 40 kHz. Maximum sensitivity for the component values shown, with VR1 fully clockwise, is 30 mV peak-to-peak. The indicator can be calibrated by setting VR1 when an appropriate input signal is applied.

LED BAR-GRAPH DRIVER



GENERAL ELECTRIC/RCA

Fig. 26-4

Circuit Notes

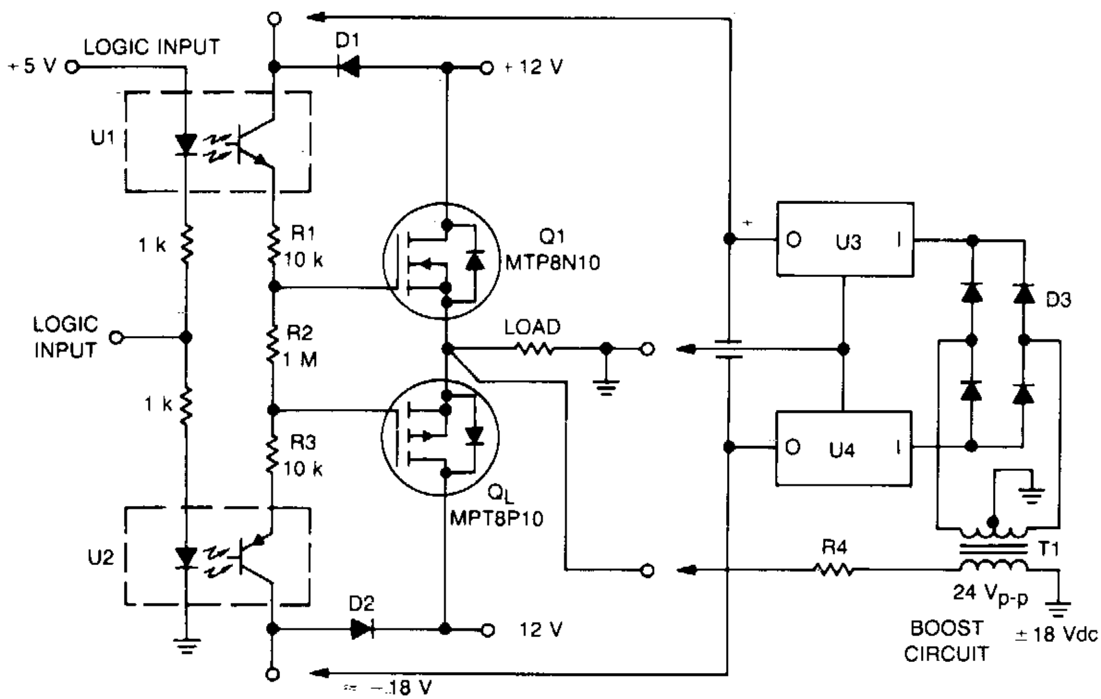
The circuit uses CA3290 BiMOS dual voltage comparators. Non-inverting inputs of A1 and A2 are tied to voltage divider reference. The input signal is applied to the inverting inputs. LEDs are turned "on" when input voltage the reaches the voltage on the reference divider.

27

Drive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LINE DRIVER PROVIDES FULL RAIL EXCURSIONS



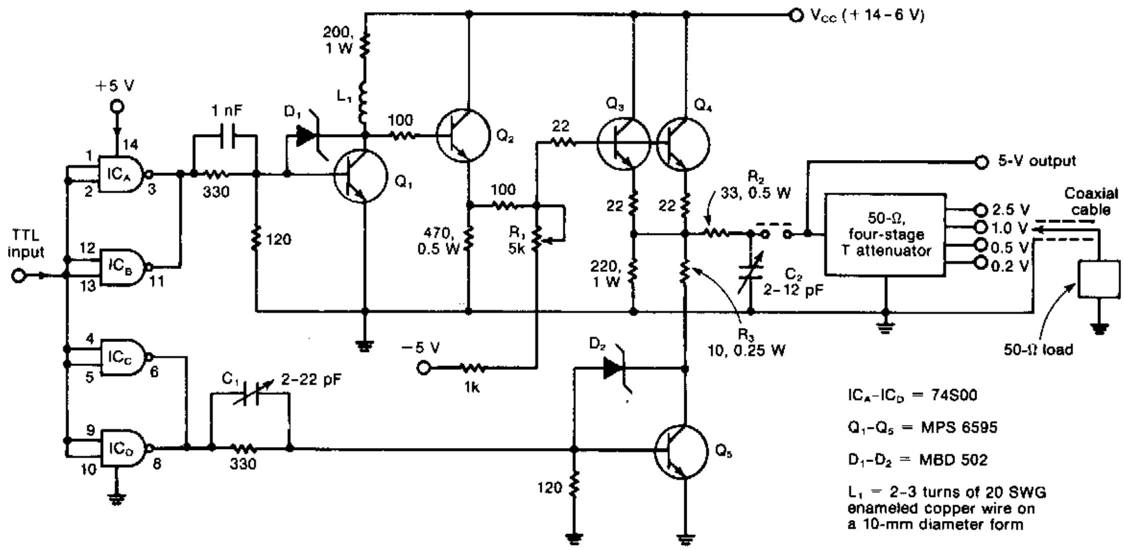
MOTOROLA *Fig. 27-1*

Circuit Notes

The logic input is applied to opto-isolators U1 and U2 with, respectively, npn and pnp emitter follower outputs. Dc balance is adjusted by potentiometer R2. The emitter followers drive the gates of Q1 and Q2, the complementary TMOS pairs. With a ±12 V supply, the swing at the common source output point is about 12 V peak-to-peak.

By adding a ±18-V boost circuit, as shown, the output swing can approach the rail swing. This circuit applies the output to transformer T1, which is rectified by diode bridge D3, regulated by U3 and U4, and then applied to the collectors of U1 and U2. Diodes D1 and D2 are forward-biased when 12-V supplies are used, but they are back-biased when the 18-V boost is used.

FIVE-TRANSISTOR AMPLIFIER BOOSTS FAST PULSES INTO 50-OHM COAXIAL CABLE



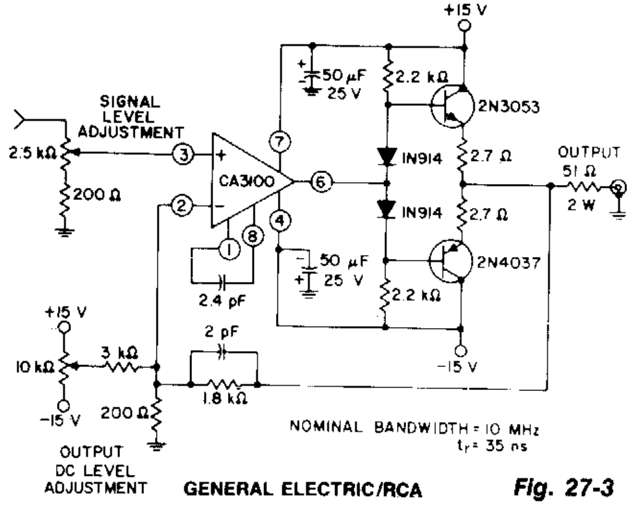
ELECTRONIC DESIGN

Fig. 27-2

Circuit Notes

The circuit works from dc to 50 MHz and will deliver pulses as short as 10 ns. It is driven by a TTL signal through a 740S00 quad Schottky NAND gate, IC_A through IC_D. Transistor Q1, wired as a common-emitter amplifier, drives transistor Q2, a simple emitter follower. Transistors Q3 and Q4, wired in parallel, also form an emitter follower and drive the output. When Q3 and Q4 are both turned off, transistor Q5 works as a low-impedance sink. Schottky diodes D1 and D2 prevent Q1 and Q5 from becoming saturated. To adjust the circuit, potentiometer R1 is set to optimize the output pulse's fall time. Inductor L1, a peaking coil, should be adjusted to improve the rise time to within a permissible 5% overshoot. Likewise, capacitor C1 can be varied to control preshooting. Further output pulse shaping is accomplished with the help of capacitor C2. Resistors R2 and R3 ensure a proper 50-ohm impedance at the amplifier's output when the pulse is on or off, respectively.

50-OHM TRANSMISSION LINE DRIVER



GENERAL ELECTRIC/RCA *Fig. 27-3*

Circuit Notes

This circuit uses a wideband, high slew rate CA3100 BiMOS op amp. The slew rate for this amplifier is 28 V/μs. Output swing is 9 volts peak-to-peak into a terminated line, measured at the termination.

600-OHM BALANCED DRIVER FOR LINE SIGNALS

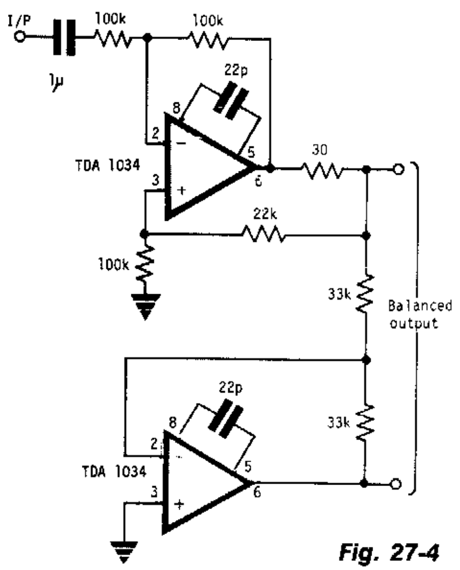


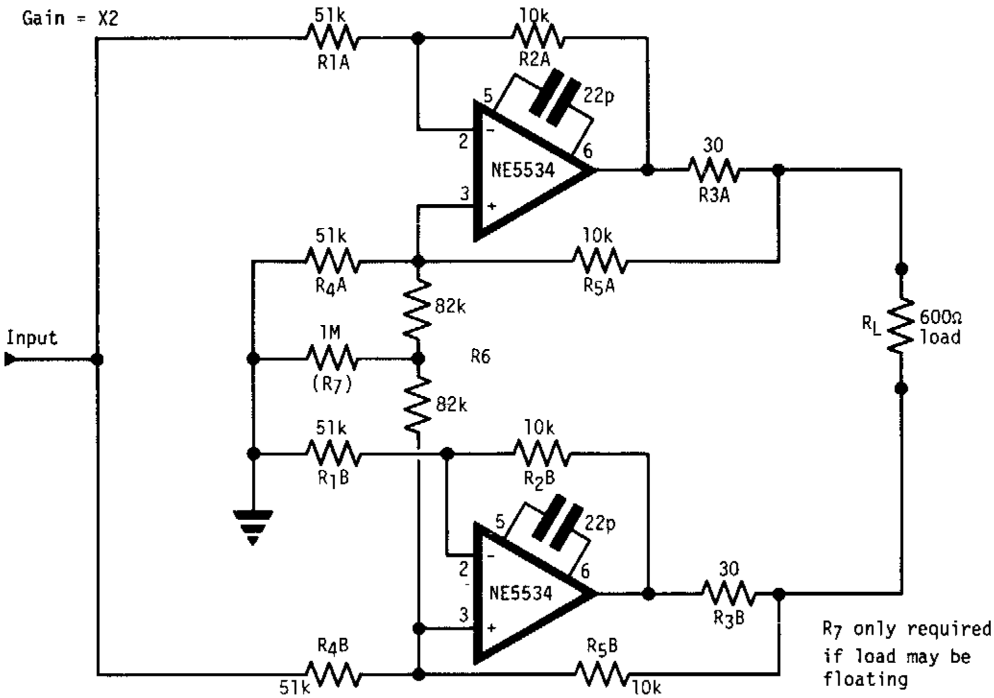
Fig. 27-4

ELECTRONIC ENGINEERING

Circuit Notes

This circuit uses current and voltage feedback. This circuit will handle +24 dBm with ±12 volts supply using TDA 1034s.

HIGH OUTPUT 600-OHM LINE DRIVER



ELECTRONIC ENGINEERING

Fig. 27-5

Circuit Notes

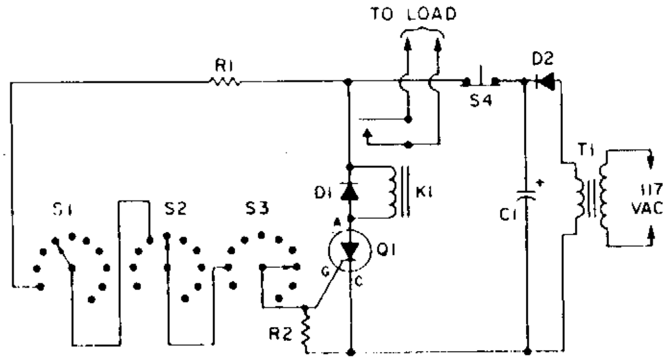
The circuit has a "floating" output, i.e., it behaves like an isolated transformer winding, with the output amplitude remaining unchanged whether the center or either end of the load is grounded. This is achieved by making Z-out, common mode, infinite. The circuit consists of two current-sources in push-pull. Since each has infinite output Z, the common mode output impedance is also infinite. Connecting a resistor between the non-inverting terminals of the op amps reduces the differential Z-out without affecting the Z-common-mode. Since the output is floating, if the load is also floating there is no output ground reference, which results in malfunction. This can be corrected by reducing the common-mode Z slightly. R7 fulfills this function. All resistors should be of close tolerance to give a good balance. The line driver provides +24 dB from ±12 V or +16 dB from ±6 V supplies.

28
Electronic Locks

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

THREE-DIAL COMBINATION LOCK

C1—500-uF, 25-VDC electrolytic capacitor
D1, D2—1N4002 diode
K1—relay with 6-volt coil rated @ 250-ohms, with SPST contacts
Q1—2N5050 SCR
R1, R2—4,700-ohm, ½-watt resistor, 5%
S1, S2, S3—single pole, 10-position rotary or thumbwheel switches
S4—normally closed SPST pushbutton switch
T1—120-VAC to 6.3-VAC @ 300mA power transformer

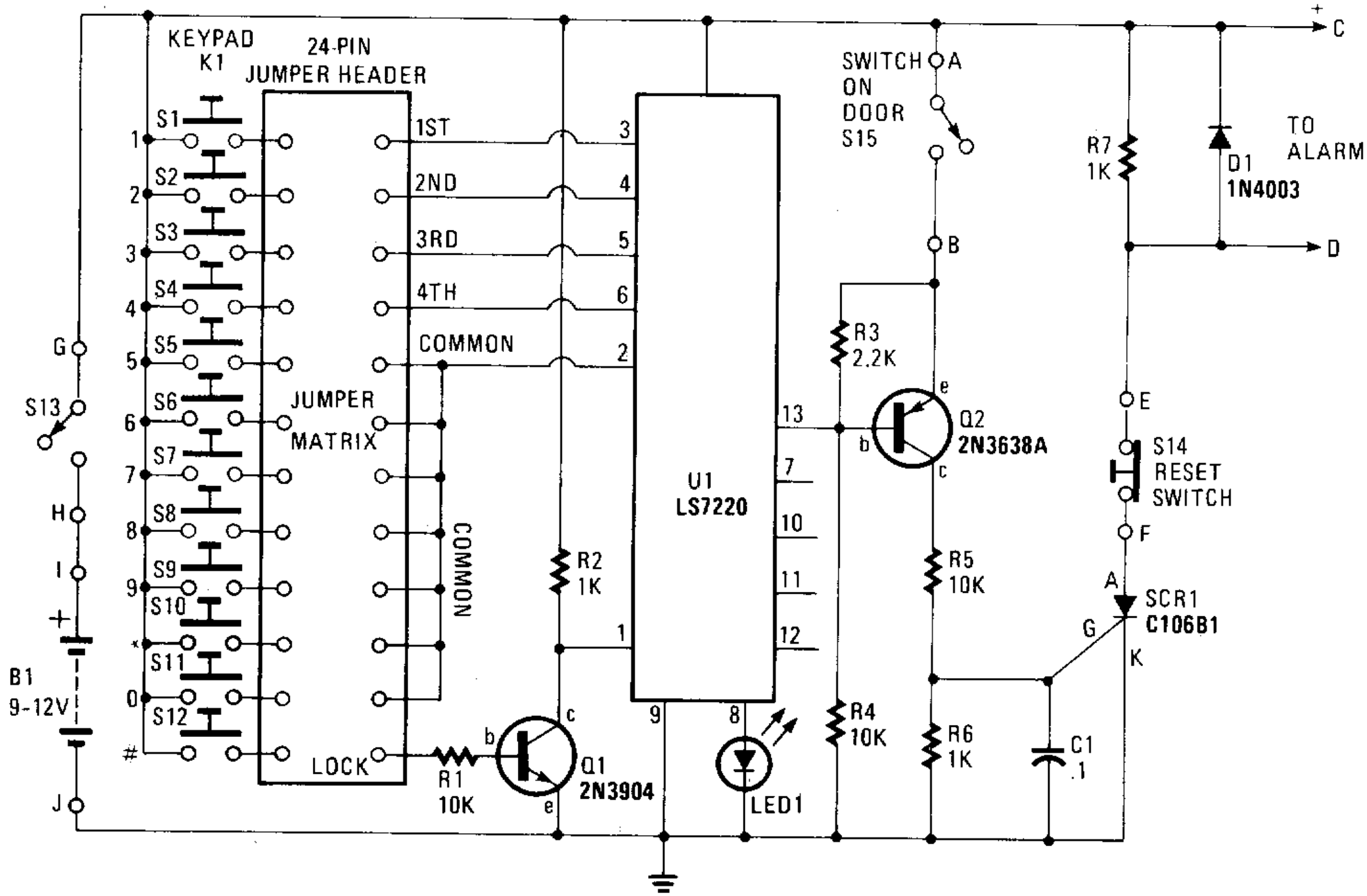


TAB BOOKS, INC. *Fig. 28-1*

Circuit Notes

Here's an effective little combination lock that you can put together in one evening's time. To open the lock, simply dial in the correct combination on the three rotary or thumbwheel switches. With the correct combination entered, current flows through R1 into Q1's gate terminal, causing the SCR to latch in a conductive state. This sends a current through relay K1, which responds by closing its contacts and actuating whatever load is attached. After opening the lock, twirl the dials of S1 through S3 away from the correct combination so that nobody gets a look at it. The lock will remain open and your load will remain on because the SCR is latched on. To lock things up, it's only necessary to interrupt the flow of anode current through the SCR by pressing pushbutton S4.

ELECTRONIC COMBINATION LOCK



HANDS-ON ELECTRONICS

Fig. 28-2

Circuit Notes

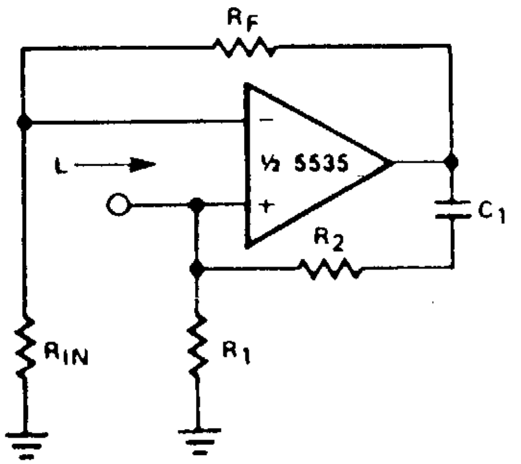
When button S12 (#) is pressed, a positive voltage fed through R1 appears at the base of transistor Q1, turning it on. When Q1 is conducting, pin 1 of U1 is brought to ground (low) or the battery's negative terminal. With pin 1 low, two things occur: Pin 8 of U1 goes high (+9 volts dc), turning on LED 1—indicating that the circuit has been armed—and pin 13 goes from high to low. Transistor Q2 requires a low signal or negative voltage on its base in order to conduct. It also needs a positive voltage on its emitter and a negative voltage on the collector. As long as the door switch (S15) remains open (with the door itself closed), Q2's emitter will not receive the necessary positive voltage. If, however, an unauthorized person opens the door, thus closing switch S15 and placing a positive voltage on the emitter of Q1, the following sequence occurs:

1. Transistor Q2 conducts, receiving the necessary biasing current through a current-divider network consisting of resistors R3 and R4.
2. As Q2 conducts, a voltage drop is developed across the voltage dividers made up of resistors R5 and R6. With R5 at 10,000 ohms and R6 at 1000 ohms, approximately one volt appears at the gate of SCR1. That's enough voltage to trigger the SCR's gate.

29
Emulator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SIMULATED INDUCTOR



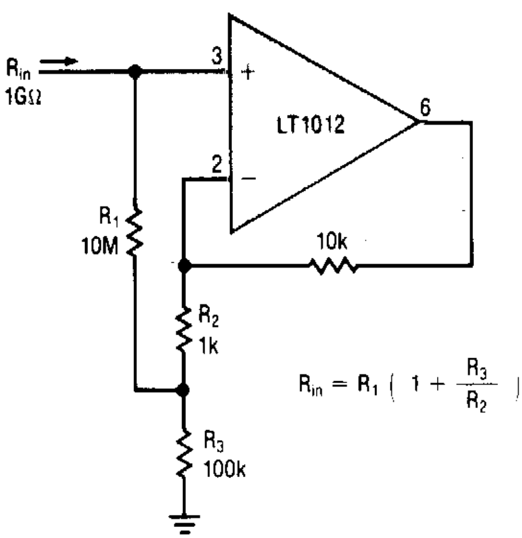
SIGNETICS

Fig. 29-1

Circuit Notes

With a constant current excitation, the voltage dropped across an inductance increases with frequency. Thus, an active device whose output increases with frequency can be characterized as an inductance. The circuit yields such a response with the effective inductance being equal to: $L = R_1R_2C$. The Q of this inductance depends upon R_1 being equal to R_2. At the same time, however, the positive and negative feedback paths of the amplifier are equal leading to the distinct possibility of instability at high frequencies. R_1 should therefore always be slightly smaller than R_2 to assure stable operation.

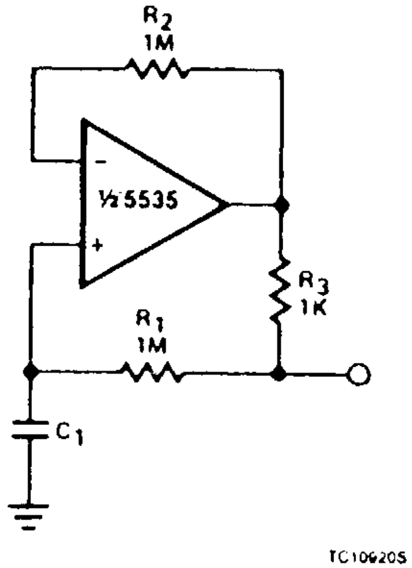
RESISTOR MULTIPLIER



LINEAR TECHNOLOGY CORP.

Fig. 29-2

CAPACITANCE MULTIPLIER



Circuit Notes

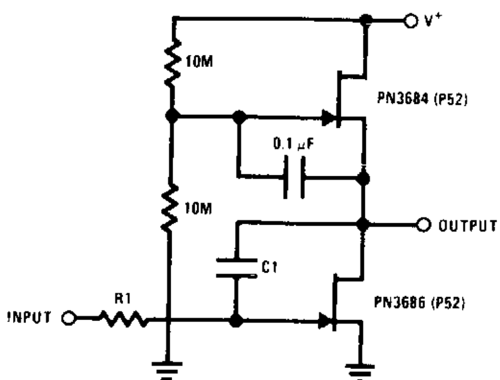
The circuit can be used to simulate large capacitances using small value components. With the values shown and C - 10 μF, an effective capacitance of 10,000 μF was obtained. The Q available is limited by the effective series resistance. So R1 should be as large as practical.

NOTE:
All resistor values are in ohms.

SIGNETICS

Fig. 29-3

JFET ac COUPLED INTEGRATOR



Circuit Notes

This circuit utilizes the "μ-amp" technique to achieve very high voltage gain. Using C1 in the circuit as a Miller integrator, or capacitance multiplier, allows this simple circuit to handle very long time constants.

NATIONAL SEMICONDUCTOR CORP.

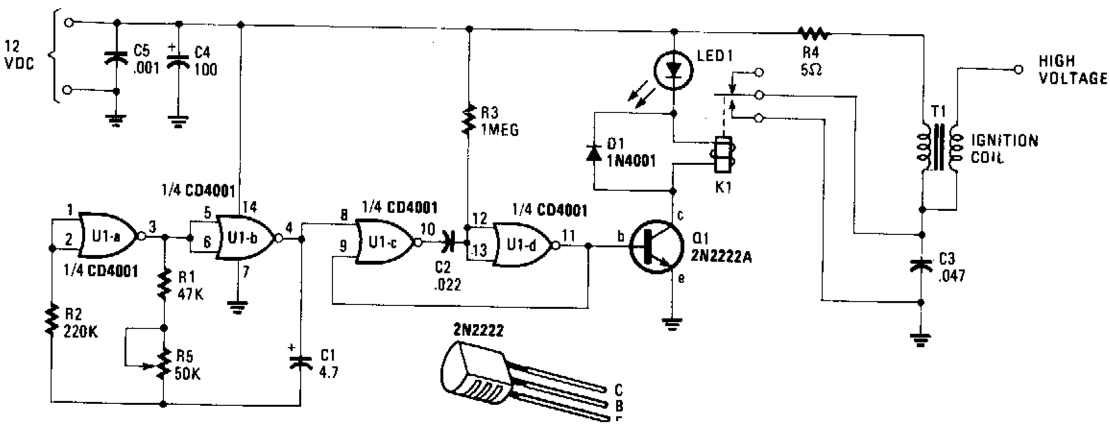
Fig. 29-4

30
Fence Chargers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Battery-Powered Fence Charger
Solid-State Electric Fence Charger
Electric Fence Charger

BATTERY-POWERED FENCE CHARGER



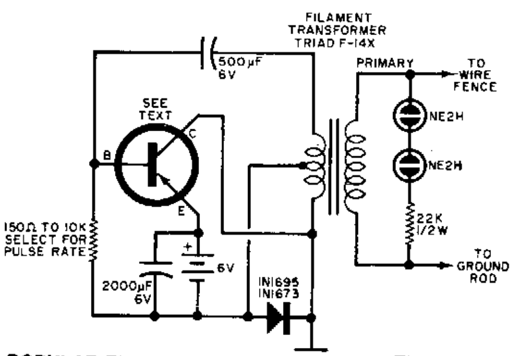
HANDS-ON ELECTRONICS

Fig. 30-1

Circuit Notes

In essence, the circuit is nothing more than an auto ignition coil and a set of points which accomplishes the same thing. A pulsing circuit made from a single CMOS NOR integrated circuit (U1), opens and closes the relay contacts to simulate the action of the original breaker points. The relay pulser is divided into two clocking functions. The first circuit is a free-running squarewave generator that determines the rate or frequency of the pulses that activate the relay. It is essentially a pair of NOR gates connected as inverters and placed in a feedback loop, they are U1-b. The oscillating period of the feedback loop is determined by timing components C1, R1, and variable resistor R5.

ELECTRIC FENCE CHARGER



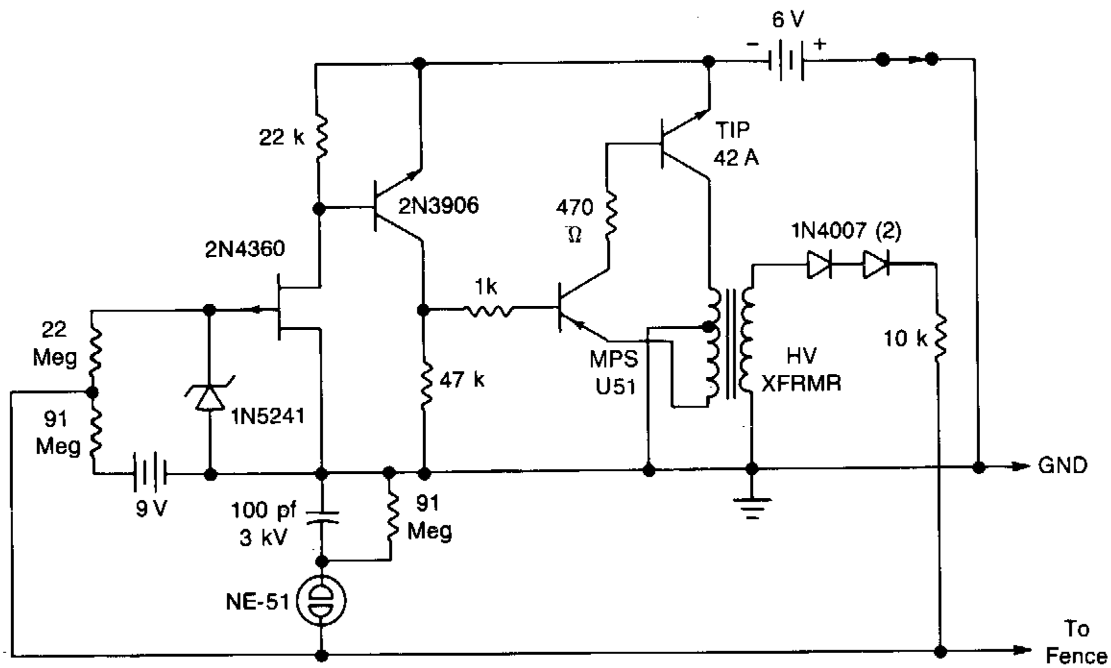
POPULAR ELECTRONICS

Fig. 30-2

Circuit Notes

Any good power transistor can be used in this circuit. The base resistor should be adjusted to obtain a pulse rate of about 50 pulses per minute. The range of values shown can go from 10 pulses to 100 pulses per minute. The single fence wire must be insulated at each supporting pole and should be mounted low enough to prevent an animal from crawling under the wire. The two neon lamps indicate when the unit is operating.

SOLID-STATE ELECTRIC FENCE CHARGER



WILLIAM SHEETS *Fig. 30-3*

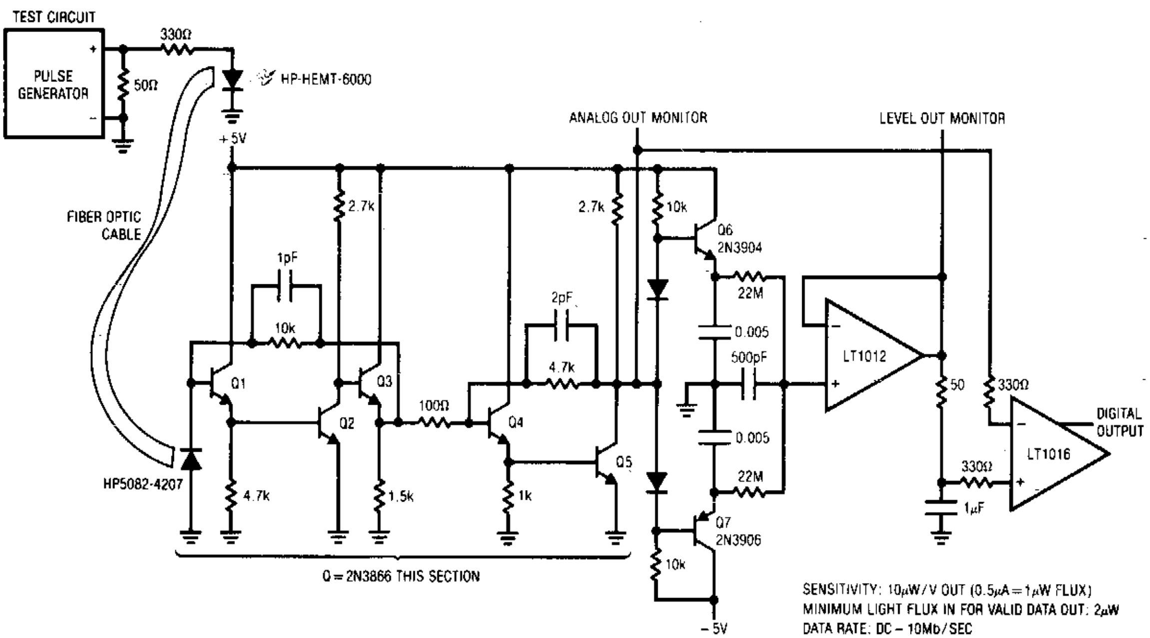
Circuit Notes

A touch-sensing circuit keeps the high-voltage generator cut off until something touches the fence wire. Contact with the fence sensing circuit wire starts the high-voltage generator which applies a series of 500 microsecond pulses at approximately 300 volts to the fence wire. Pulse repetition rate is determined by the intruder's resistance to earth ground. The lower the resistance, the higher the pulse rate. A ground rod is inserted several inches into the ground near the fence wire. In the sensing mode the neon lamp should not flicker or light. If it does, it indicates leakage between the fence wire and ground. If sensitivity is too great, it can be reduced by changing the 91 Meg resistor to 47 or 22 Meg as required.

31
Fiberoptics Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

10 MHz FIBEROPTIC RECEIVER

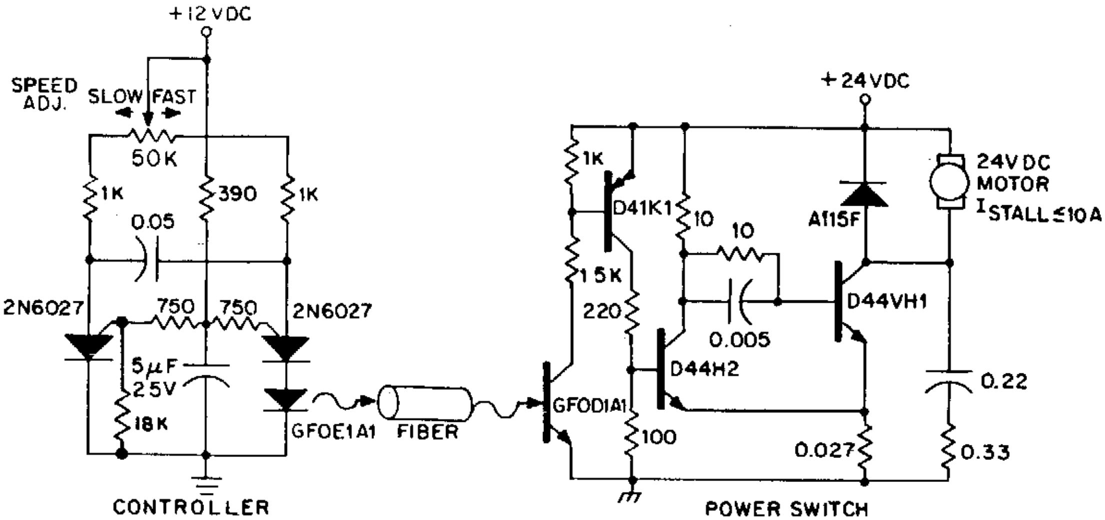


LINEAR TECHNOLOGY CORPORATION *Fig. 31-1*

Circuit Notes

The receiver will accurately condition a wide range of light inputs at up to 10 MHz data rates. The optical signal is detected by the PIN photodiode and amplified by a broadband fed-back stage, Q1-Q3. A second, similar, stage gives further amplification. The output of this stage (Q5's collector) biases a 2-way peak detector (Q6-Q7). The maximum peak is stored in Q6's emitter capacitor while the minimum excursion is retained in Q7's emitter capacitor. The dc value of Q5's output signal's mid-point appears at the junction of the 0.005 μF capacitor and the 22 M ohm unit. This point will always sit midway between the signal's excursions, regardless of absolute amplitude. This signal-adaptive voltage is buffered by the low bias LT1012 to set the trigger voltage at the LT1016's positive input. The LT1016's negative input is biased directly from Q5's collector.

DC VARIABLE SPEED MOTOR CONTROL VIA FIBEROPTICS



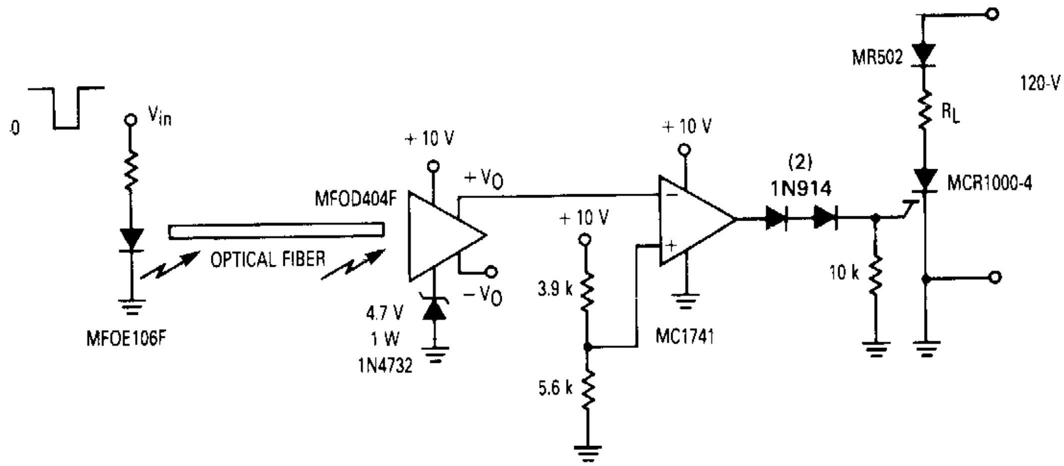
GENERAL ELECTRIC

Fig. 31-2

Circuit Notes

Dc power can also be controlled via fiberoptics. The circuit provides an insulated speed control path for a small dc actuator motor (≤ 1/12 hp). Control logic is a self-contained module requiring about 300 mW at 12 V, which can be battery powered. The control module furnishes infrared pulses, at a rate of 160 Hz, with a duty cycle determined by the position of the speed adjust potentiometer. The programmable unijunction multivibrator provides approximately 10 mA pulses to the GFOE1A1 at duty cycles adjustable over a range of 1% to 99%. The infrared pulses are detected by the GFOD1A1, amplified by the D39C1 pnp Darlington, and supplied to the power drive switch, which is connected in a Schmitt trigger configuration to supply the motor voltage pulses during the infrared pulses. Thus, the motor's average supply voltage is pulse width modulated to the desired speed, while its current is maintained between pulses by the A115F free-wheeling diode. The snubber network connected in parallel with the power switch minimizes peak power dissipation in the output transistor, and enhancing reliability. Larger hp motors can be driven by adding another stage of current gain, while longer fiber range lengths can be obtained with an amplifier transistor driving the GFOE1A1.

FIBEROPTIC INTERFACE



MOTOROLA

Fig. 31-3

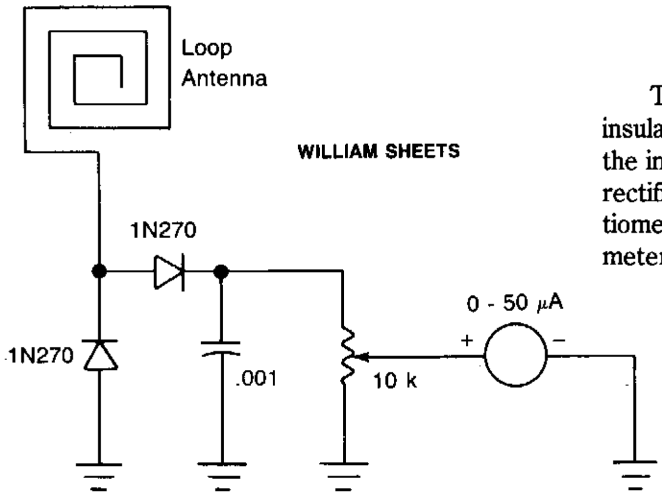
Circuit Notes

An op amp is used to interface between a fiberoptic system and the MOS SCR to multi-cycle, half-wave control of a load. This receiver has two complementary outputs, one at a quiescent level of about 0.6 V and the second at 3 V. By adding a 4.7 V zener in series with the return bus, the effective V_{CC} becomes 5.3 V and also the 0.6 V output level is translated up to about 5.3 V. This level is compatible with the reference input (5.9 V) of the single-ended powered op-amp acting as a comparator.

32
Field Strength Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FIELD-STRENGTH METER



Circuit Notes

The antenna consists of about 20 cm of insulated stranded wire glued or taped around the inside of a small plastic box. RF current is rectified by two diodes, and a 10 k potentiometer provides variable attenuation for the meter.

Fig. 32-1

FIELD-STRENGTH METER II

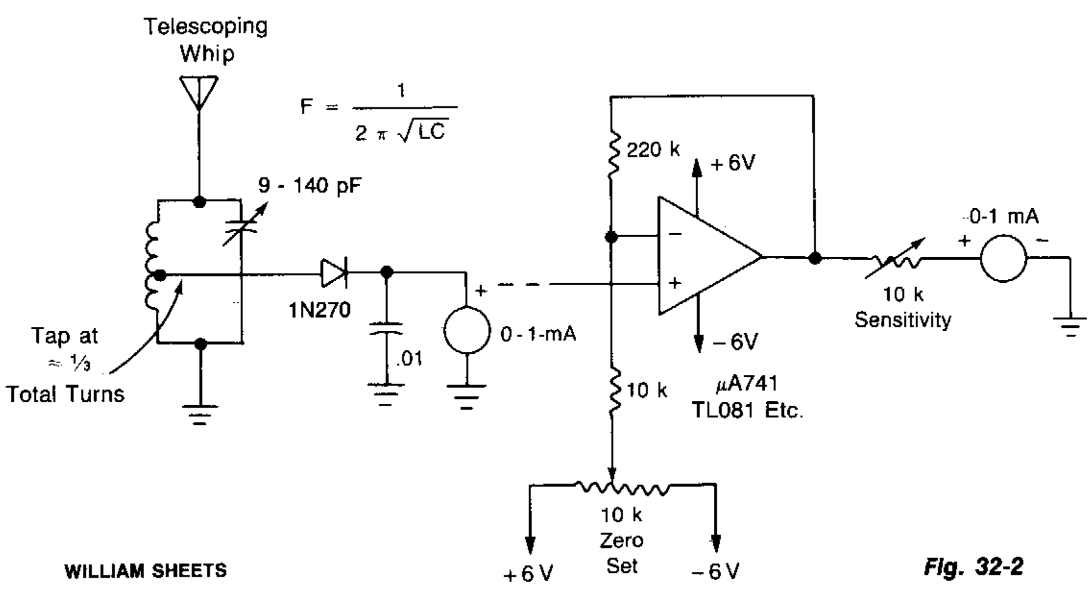
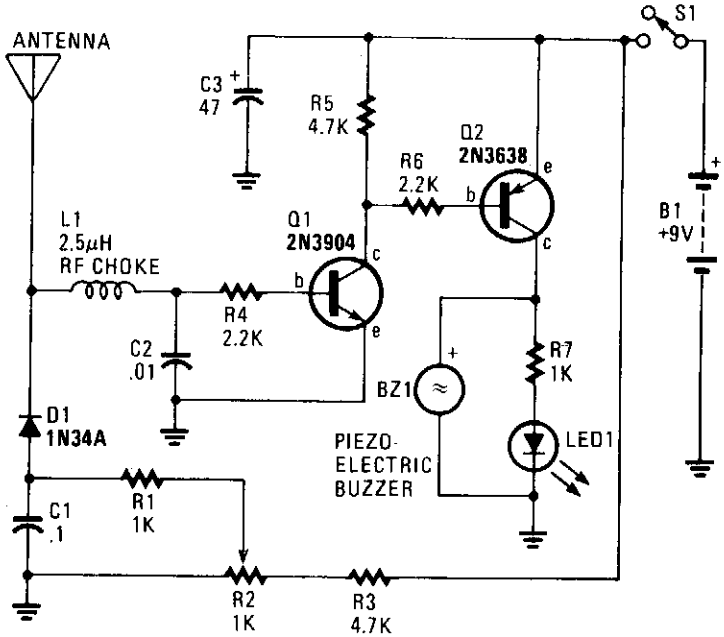


Fig. 32-2

Circuit Notes

"Minimum-parts" field-strength meter is shown here. For more distant testing, add the dc amplifier.

RF SNIFFER



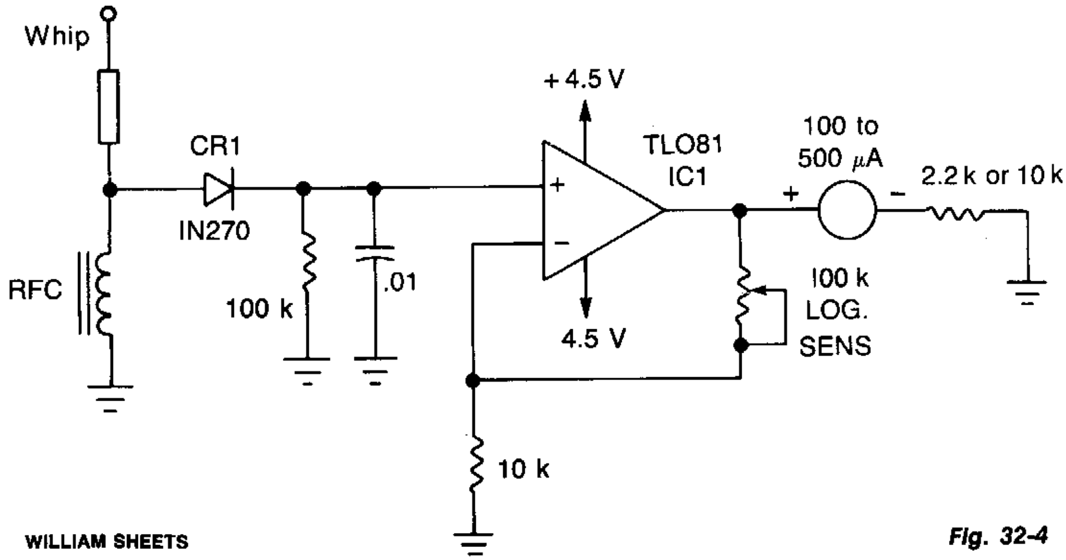
HANDS-ON ELECTRONICS

Fig. 32-3

Circuit Notes

This circuit responds to RF signals from below the standard broadcast band to well over 500 MHz, and provides a visual and audible indication when a signal is received. The circuit is designed to receive low-powered signals as well as strong sources of energy by adjusting the bias on the pick-up diode, D1, with R2. A very sensitive setting can be obtained by carefully adjusting R2 until the LED just begins to light and a faint sound is produced by the Piezo sounder.

HIGH-SENSITIVITY FIELD STRENGTH METER



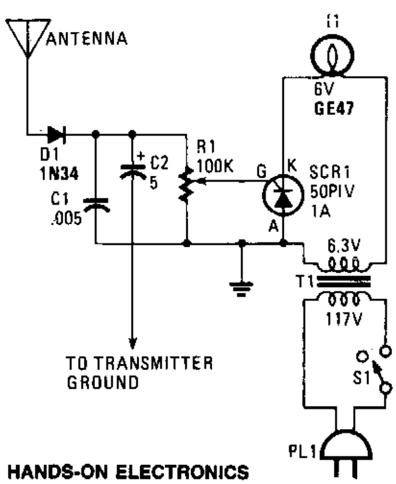
WILLIAM SHEETS

Fig. 32-4

Circuit Notes

A TL081 (IC1 op amp is used to increase sensitivity. RF signal is detected by CR1 and is then amplified by IC1. Full-scale sensitivity is set with the 100 K potentiometer.

TRANSMISSION INDICATOR



HANDS-ON ELECTRONICS

Fig. 32-5

Circuit Notes

Everytime the push-to-talk button is closed the light will go on. The antenna samples the output RF from the transmitter. That signal is then rectified (detected) by germanium diode D1, and used to charge capacitor C2. The dc output is used to trigger a small silicon-controlled rectifier (SCR1), which permits the current to flow through the small pilot lamp. For lower-power applications, such as CB radio, the antenna will have to be close-coupled to the antenna.

LF OR HF FIELD STRENGTH METER

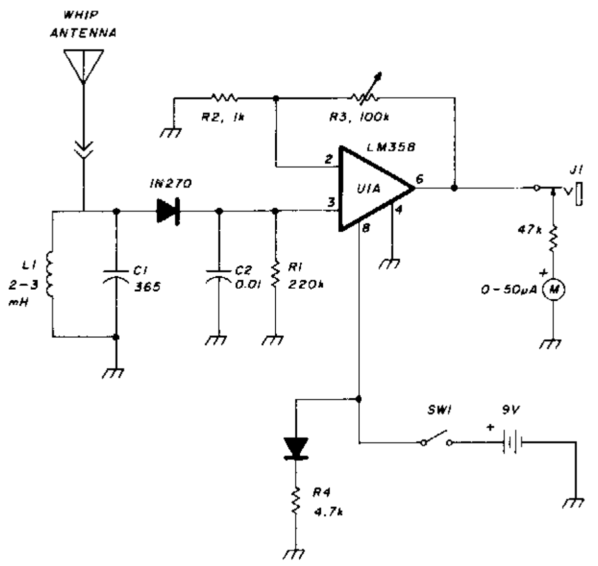


Table 1.

L1	C1 (variable)	Frequency Range	Ham Band
50 μH	30-365 pF	1- 4 MHz	160, 80 meters
3 μH	30-365 pF	5-16 MHz	40, 30, 20 meters
0.9 μH	30-365 pF	9-30 MHz	30, 20, 15, 12, 10 meters
2.5 mH	—	Broadband at reduced gain	

HAM RADIO

Fig. 32-6

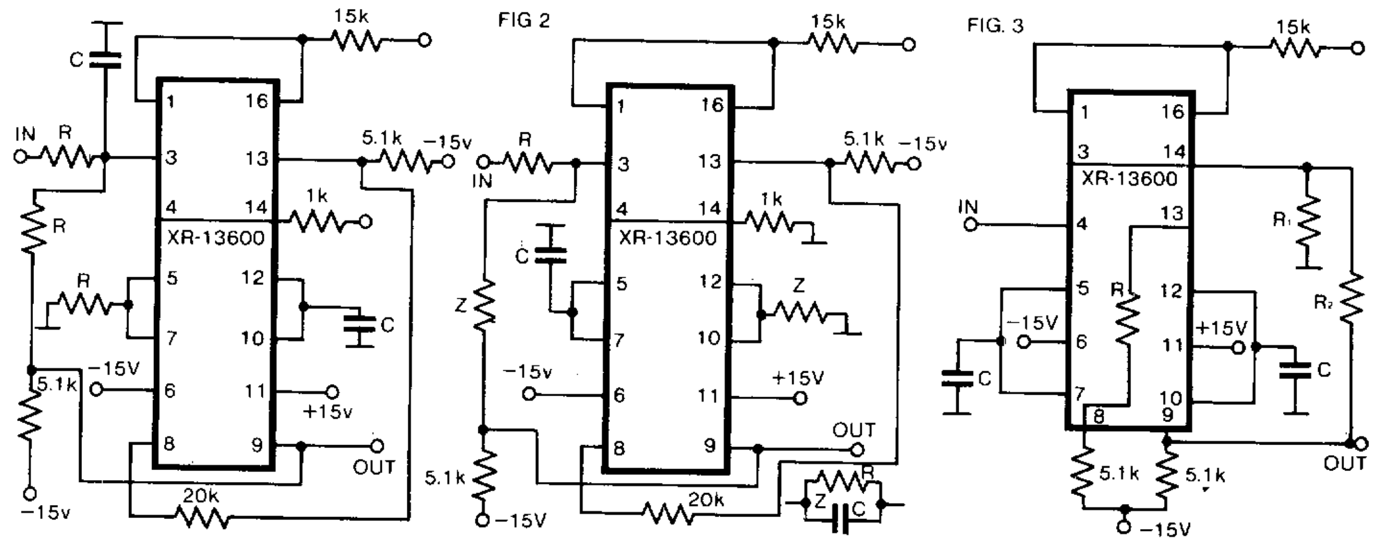
Circuit Notes

C1 and L1 resonate on the 1750 meter band, with coverage from 150 kHz to 500 kHz. L1 can be slug-tuned for 160-to-190 kHz coverage alone or a 2.5 mH choke can be used for L1, if desired, using C1 for tuning. A 1N270 germanium diode rectifies the RF signal and C2 is charged at the peak RF level. This dc level is amplified by an LM358. The gain is determined by R2 and R3, 1 100-kilohm linear potentiometer that varies the dc gain from 1 to 100, driving the 50 microampere meter. This field strength meter need not be limited to LF use. The Table shows the L1 and C1 values for HF operation and broadband operation.

33
Filter Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW COST UNIVERSAL ACTIVE FILTER



ELECTRONIC ENGINEERING

Fig. 33-1

Circuit Notes

The circuit as shown in Fig. 1 gives the bandpass operation the transfer function calculated from

$$F_{BP}(s) = \frac{S/\omega_o}{K}$$

where $K = 1 + s/Q\omega_o + s^2/\omega_o^2$.
The cut-off frequency, ω_o, and the Q-factor are given by

$$\omega_o = g/C \text{ and } Q = gR/2$$

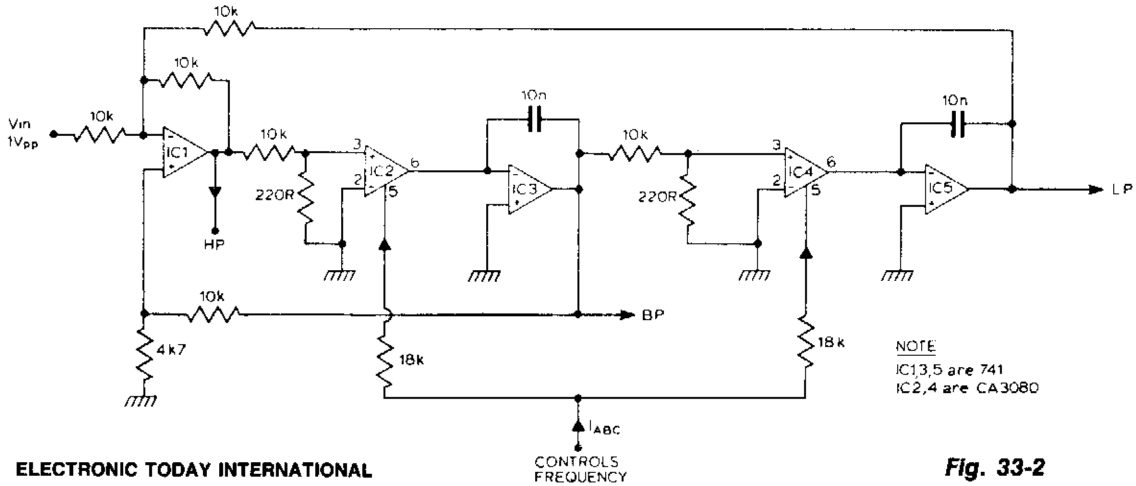
where g is the transconductance at room temperature.

Interchanging the capacitor C with the resistor R at the input of the circuit high-pass operation is obtained. A low-pass filter is obtained by applying two parallel connections of R and C as shown in Fig. 2.

The low-pass operation may be much improved with the circuit as given in Fig. 3. Here the gain and Q may be set up separately with respect to the cut-off frequency according to the equations

$$Q = 1/fB = 1 + R_2/R_1,$$
$$A = Q^2 \text{ and } \omega_o = g\, fB/C.$$

STATE VARIABLE FILTER



ELECTRONIC TODAY INTERNATIONAL

Fig. 33-2

Circuit Notes

The filter produces three outputs: high-pass, bandpass, and low-pass. Frequency is linearly proportional to the gain of the two integrators. Two CA3080's, (IC2, 4) provide the variable gain, the resonant frequency being proportional to the current I_{ABC}. Using 741 op amps for IC3 a control range of 100 to 1, (resonant frequency) can be obtained. If CA3140's are used instead of 741's then this range can be extended to nearly 10,000 to 1.

WIDEBAND TWO-POLE HIGH-PASS FILTER

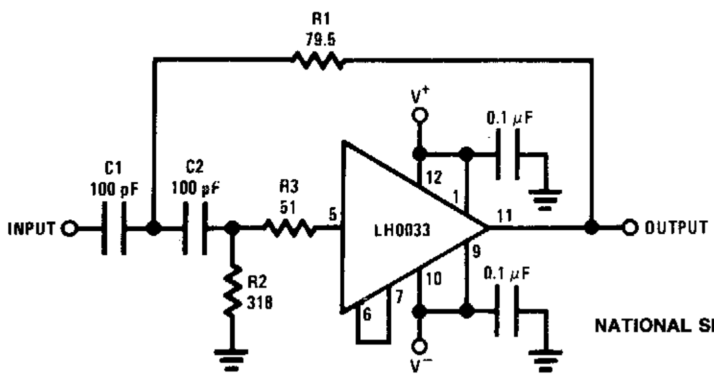


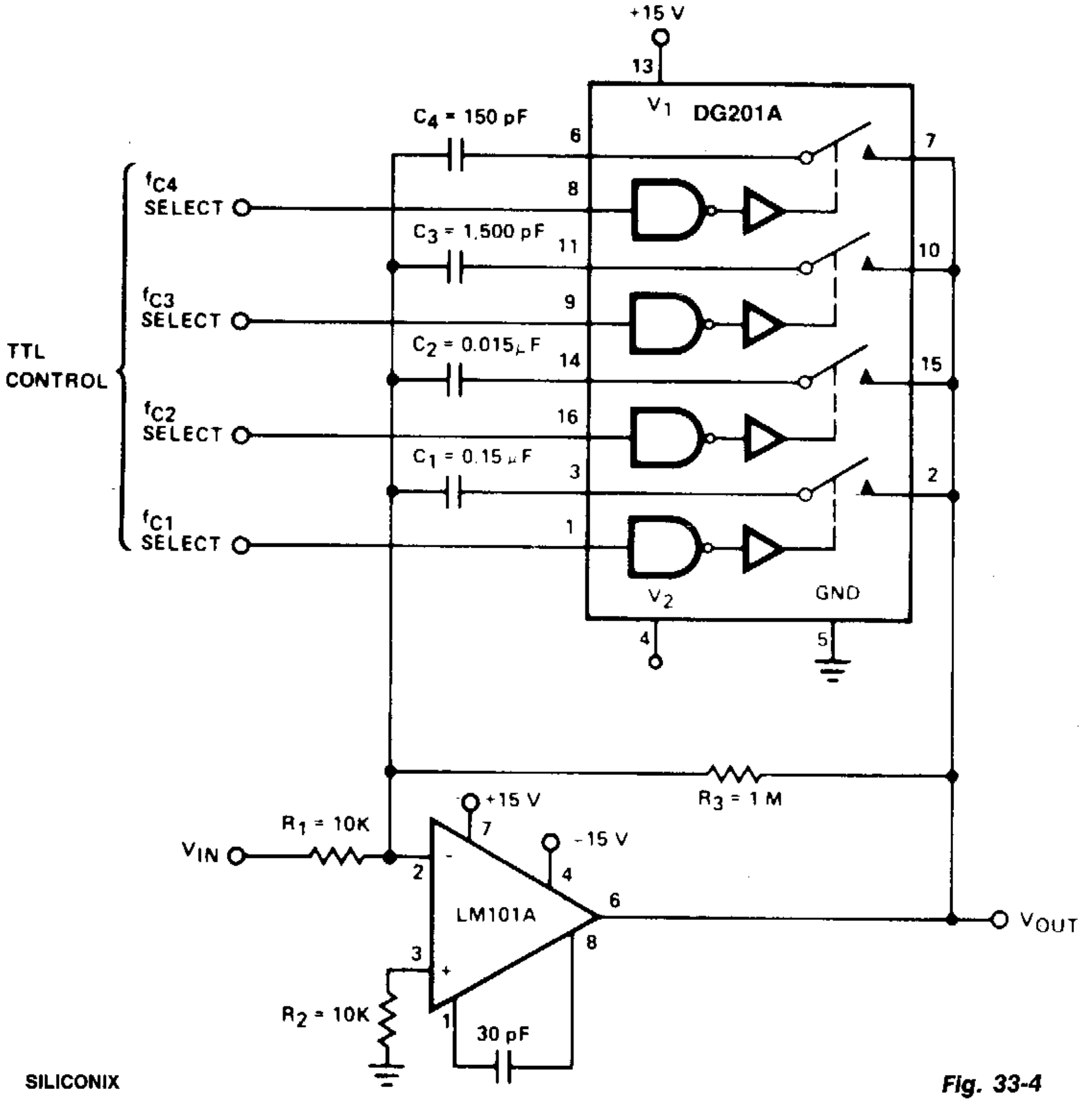
Fig. 33-3

NATIONAL SEMICONDUCTOR CORP.

Circuit Notes

The circuit provides a 10MHz cutoff frequency. Resistor R3 ensures that the input capacitance of the amplifier does not interact with the filter response at the frequency of interest. An equivalent low pass filter is similarly obtained by capacitance and resistance transformation.

ACTIVE LOW-PASS FILTER WITH DIGITALLY SELECTED BREAK FREQUENCY



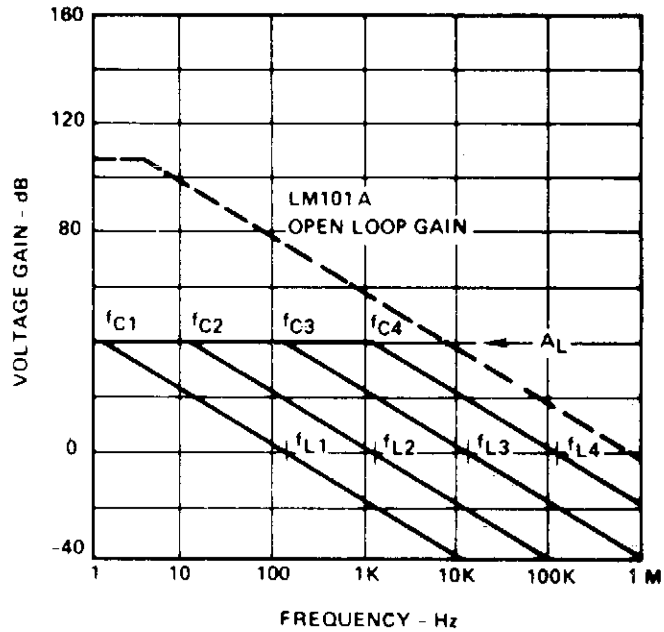
SILICONIX

Fig. 33-4

Circuit Notes

Variable low-pass filter has break frequencies at 1, 10, 100 Hz and 1 kHz. The break frequency is

$$1.\quad fc = \frac{1}{2\pi R_3 C_X}$$



A_L (VOLTAGE GAIN BELOW BREAK FREQUENCY)

$= \frac{R_3}{R_1} = 100$ (40 dB)

f_c (BREAK FREQUENCY) $= \frac{1}{2\pi R_3 C_X}$

f_L (UNITY GAIN FREQUENCY) $= \frac{1}{2\pi R_1 C_X}$

MAX ATTENUATION $= \frac{r_{DS(on)}}{10K} \approx -40$ dB

The low frequency gain is

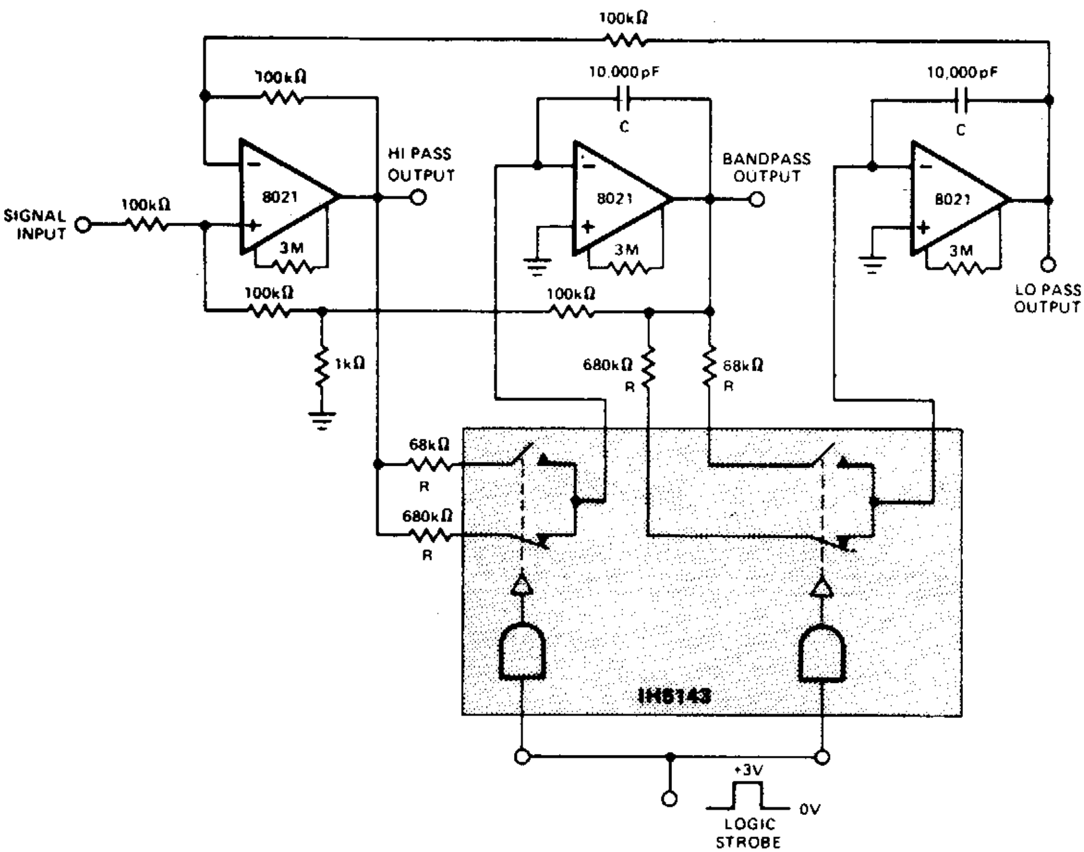
$$2. \quad A_L = \frac{R_3}{R_3} = 100 \text{ (40 dB)}$$

A second break frequency (a zero) is introduced by $r_{DS(on)}$ of the DG201A, causing the minimum gain to be

$$3. \quad A_{MIN} = \frac{r_{DS(on)}}{R_1} \approx \frac{100}{10K} = .01,$$

a maximum attenuation of 40 dB (80 dB relative to the low frequency gain).

DIGITALLY TUNED LOW POWER ACTIVE FILTER



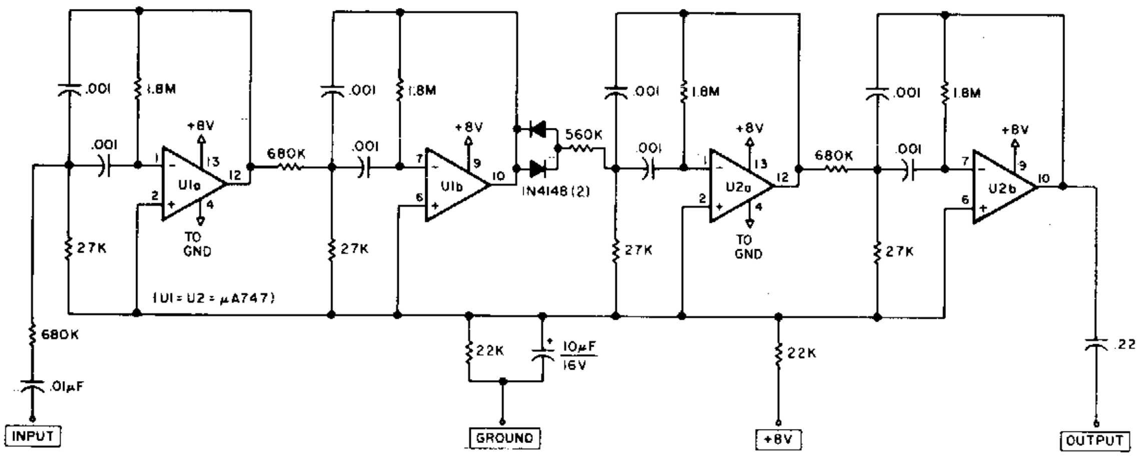
INTERSIL

Fig. 33-5

Circuit Notes

This constant gain, constant Q, variable frequency filter provides simultaneous low-pass, bandpass, and high-pass outputs with the component values shown, the center frequency will be 235 Hz and 23.5 Hz for high and low logic inputs respectively, Q = 100, and gain = 100.

RAZOR-SHARP CW FILTER



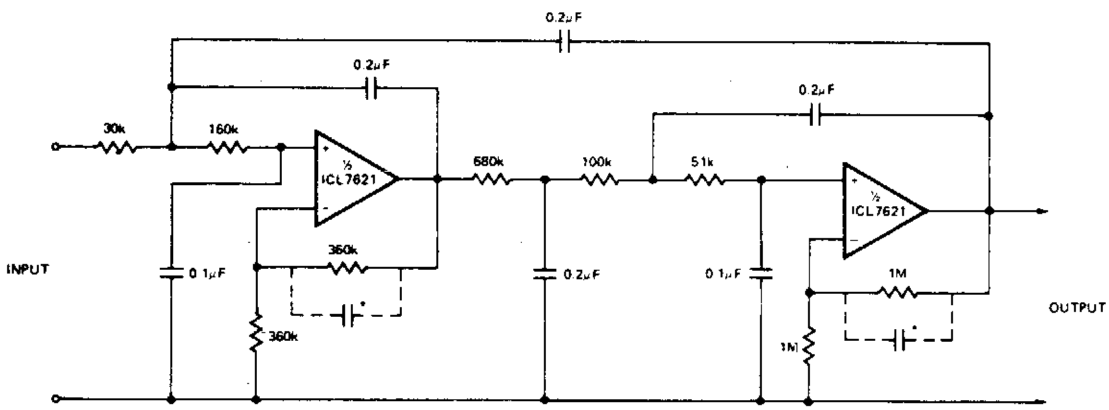
73 MAGAZINE

Fig. 33-6

Circuit Notes

The circuit consists of four stages of active bandpass filtering provided by two type-μA747 integrated-circuit dual op amps and includes a simple threshold detector (diodes D1 and D2) between stages 2 and 3 to reduce low-level background noise. Each of the four filter stages acts as a narrow bandpass filter with an audio bandpass centered at 750 Hz. The actual measured 3-dB bandwidth is only 80 Hz wide.

FIFTH ORDER CHEBYSHEV MULTIPLE FEEDBACK LOW PASS FILTER



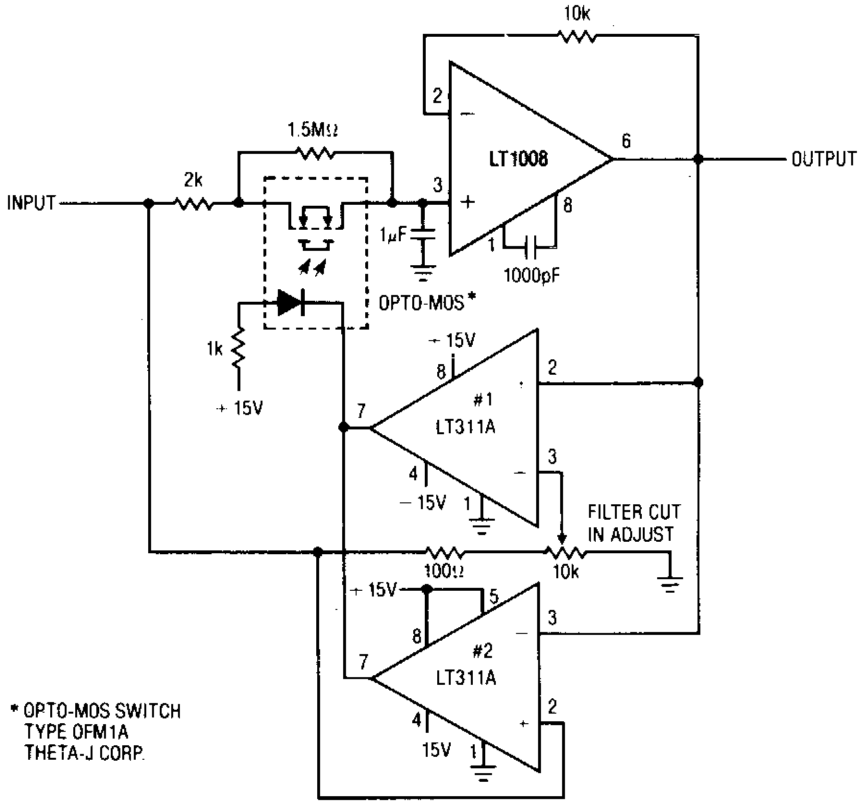
INTERSIL

Fig. 33-7

Circuit Notes

The low bias currents permit high resistance and low capacitance values to be used to achieve low frequency cutoff. f_c = 10 Hz, A_{VCL} = 4, Passband ripple = 0.1 dB. Note that small capacitors (25-50 pF) may be needed for stability in some cases.

PRECISION, FAST SETTLING, LOW-PASS FILTER

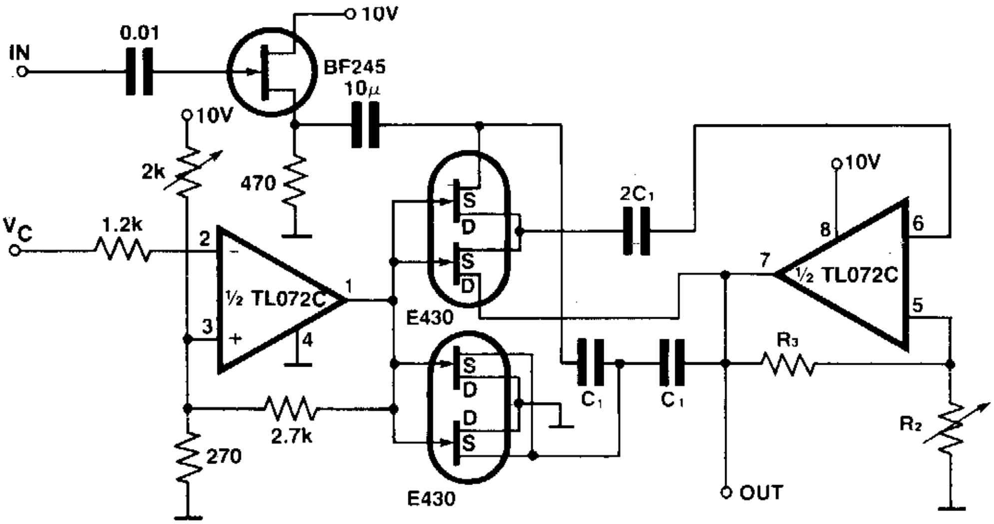


LINEAR TECHNOLOGY ***Fig. 33-8***

Circuit Notes

This circuit is useful where fast signal acquisition and high precision are required, as in electronic scales. The filter's time constant is set by the 2 K ohm resistor and the 1 μF capacitor until comparator No. 1 switches. The time constant is then set by the 1.5 M ohm resistor and the 1 μF capacitor. Comparator No. 2 provides a quick reset. The circuit settles to a final value three times as fast as a simple 1.5 M ohm—1 μF filter, with almost no dc error.

PROGRAMMABLE BANDPASS USING TWIN-T BRIDGE



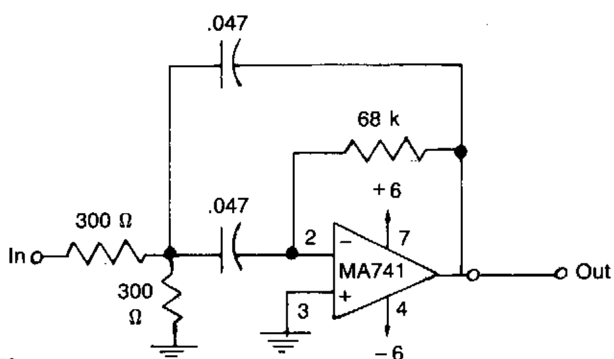
ELECTRONIC ENGINEERING

Fig. 33-9

Circuit Notes

The circuit gives a programmable bandpass where both the cut-over frequency and the gain, A, are controlled independently. In the twin-T bridge the resistors R and R/2 are replaced by two double FETs, E 430, the channel resistance of the first one in the series, the channel resistances of the second one are in parallel as to stimulate the resistance R/2. Both these resistors are controlled by V_c which ranges from 0 V to about 1 V. The gain of the circuit is set by means of the resistors R2 and R3.

ACTIVE BANDPASS FILTER (f0 = 1000 Hz)



WILLIAM SHEETS

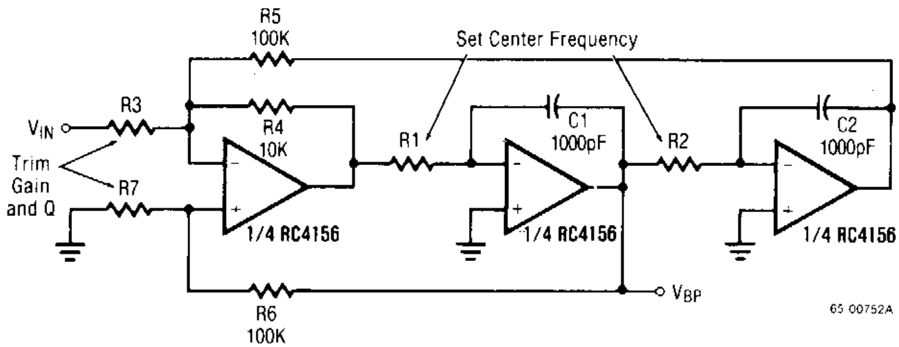
Fig. 33-10

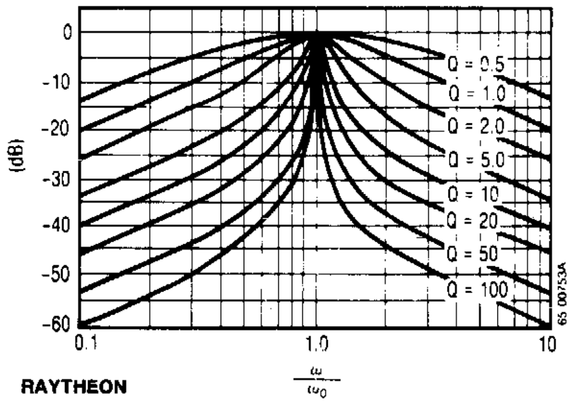
Circuit Notes

This filter has a bandpass centered around 1kHz, for applications such as bridge amplifiers, null detectors, etc.

The circuit uses a μA741 IC and standard 5% tolerance components.

BANDPASS FILTER





$$\frac{V_{BP}}{V_{IN}} = \frac{\frac{\omega}{\omega_0}\ \frac{1}{Q}}{\sqrt{\left[1-\left(\frac{\omega}{\omega_0}\right)^2\right]^2 + \left(\frac{1}{Q}\ \frac{\omega}{\omega_0}\right)^2}}$$

RAYTHEON ***Fig. 33-11***

Circuit Notes

The input signal is applied through R3 to the inverting input of the summing amplifier and the output is taken from the first integrator. The summing amplifier will maintain equal voltage at the inverting and non-inverting inputs. Defining 1/R1C1 as ω_1 and 1/R2C2 as ω_2, this is now a convenient form to look at the center-frequency ω_0 and filter Q.

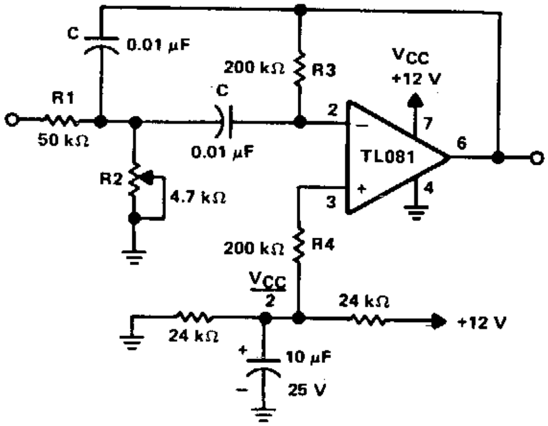
$$\omega_0 = \sqrt{0.1\ \omega_1\ \omega_2}$$

$$\text{and } Q = \left[\frac{1 + \frac{10^5}{R7}}{1.1 + \frac{10^4}{R3}}\right] \omega_0$$

$$= 10^{-9}\ \sqrt{0.1 R1 R2}$$

The frequency response for various values of Q is shown.

ACTIVE BANDPASS FILTER



TEXAS INSTRUMENTS

Fig. 33-12

Circuit Notes

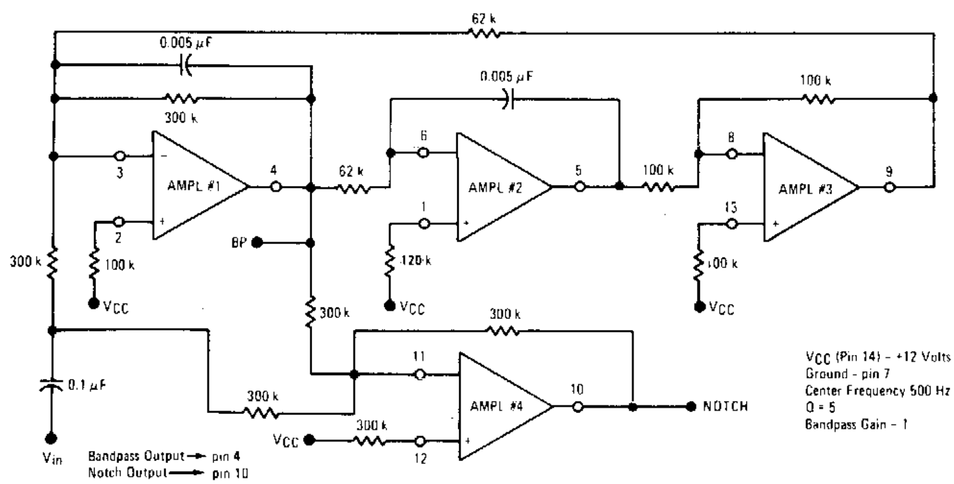
The circuit is a two-pole active filter using a TL081 op amp. This type of circuit is usable only for Qs less than 10. The component values for this filter are calculated from the following equations.

$$R1 = \frac{Q}{2\ fGC} \qquad R3 = \frac{2Q}{2fC}$$

$$R2 = \frac{Q}{(2Q^2 - G)2fC} \qquad R4 = R3$$

The values shown are for a center frequency of 800 Hz.

BANDPASS AND NOTCH FILTER



MOTOROLA

Fig. 33-13

Circuit Notes

The Quad op amp MC4301 is used to configure a filter that will notch out a given frequency and produce that notched-out frequency at the BP terminal, useful in communications or measurement setups. By proper component selection any frequency filter up to a few tens of kilohertz can be obtained.

MULTIPLE-FEEDBACK BANDPASS FILTER

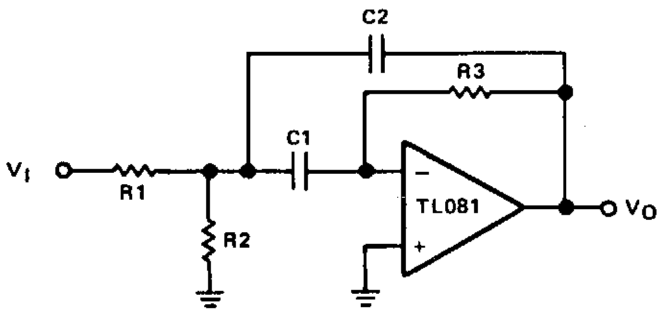


Fig. 33-14

TEXAS INSTRUMENTS

Circuit Notes

The op amp is connected in the inverting mode. Resistor R3 from the output to the inverting input sets the gain and current through the frequency-determining capacitor, C1. Capacitor C2 provides feedback from the output to the junction of R1 and R2. C1 and C2 are always equal in value. Resistor R2 may be made adjustable in order to adjust the center frequency which is determined from:

$$fo = \frac{1}{2\pi C} \quad \frac{1}{R3} \times \frac{R1 + R2^{1/2}}{R1R2}$$

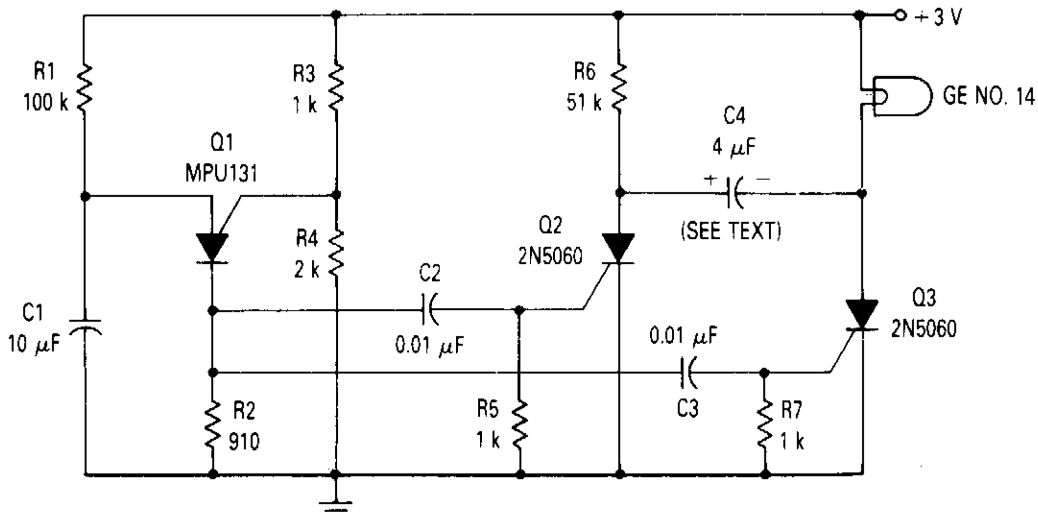
When designing a filter of this type it is best to select a value for C1 and C2, keeping them equal. Typical audio filters have capacitor values from 0.01 μF to 0.1 μF which will result in reasonable values for the resistors.

34

Flashers and Blinkers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW VOLTAGE LAMP FLASHER



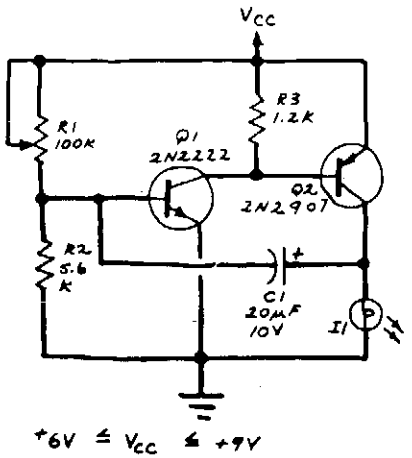
MOTOROLA

Fig. 34-1

Circuit Notes

The circuit is composed of a relaxation oscillator formed by Q1 and an SCR flip-flop formed by Q2 and Q3. With the supply voltage applied to the circuit, the timing capacitor C1 charges to the firing point of the PUT, 2 volts plus a diode drop. The output of the PUT is coupled through two 0.02 μF capacitors to the gate of Q2 and Q3. To clarify operation, assume that Q3 is on and capacitor C4 is charged plus to minus as shown in the figure. The next pulse from the PUT oscillator turns Q2 on. This places the voltage on C4 across Q3 which momentarily reverse biases Q3. This reverse voltage turns Q3 off. After discharging, C4 then charges with its polarity reversed to that shown. The next pulse from Q1 turns Q3 on and Q2 off. Note that C4 is a non-polarized capacitor. For the component values shown, the lamp is on for about ½ second and off the same amount of time.

MINIATURE TRANSISTORIZED LIGHT FLASHER



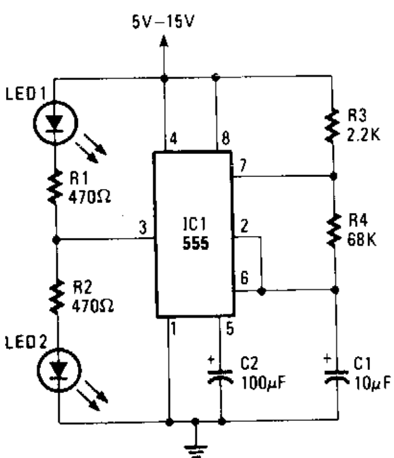
POPULAR ELECTRONICS

Fig. 34-2

Circuit Notes

R1 adjusts the flash rate. The lamp should be a No. 122, No. 222 or other similar, miniature incandescent lamp.

ALTERNATING FLASHER



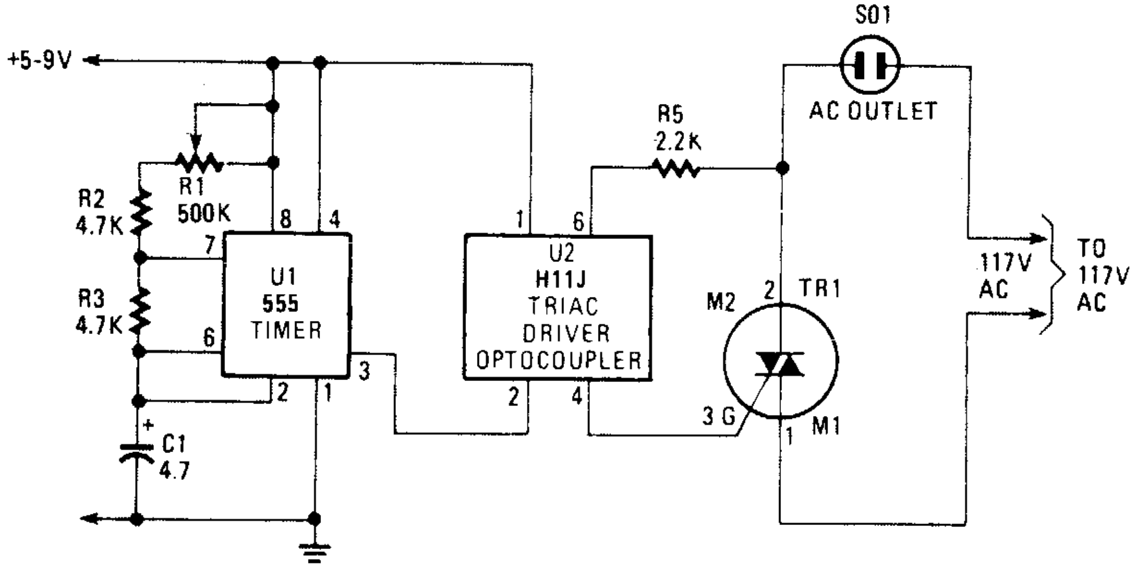
HANDS-ON ELECTRONICS

Fig. 34-3

Circuit Notes

The LED's flash alternately. The flash rate is determined by C1 and R4.

ELECTRONIC LIGHT FLASHER



HANDS-ON ELECTRONICS

Fig. 34-4

Circuit Notes

The blinking or flashing rate is determined by U1, a 555 timer integrated circuit. Its output, at pin 3, feeds U2, a HIIJ triac driver. That driver consists of an infrared LED that is coupled internally to a light-activated silicon bilateral switch (DIAC). When the LED internal to U2 is turned on by the timer, U1, its light triggers the DIAC; effectively closing the circuit between pins 4 and 6, and fires the Triac, TR1 through its gate circuit. When the Triac is firing, it acts as a closed circuit that turns on the light (or other device it may be controlling via S01). When the timer turns off, the LED, the DIAC and Triac stop conducting and the light turns off. The sequence then repeats. The flashing rate can be varied by means of R1, a 500,000 ohm potentiometer.

HIGH-POWER BATTERY-OPERATED FLASHER

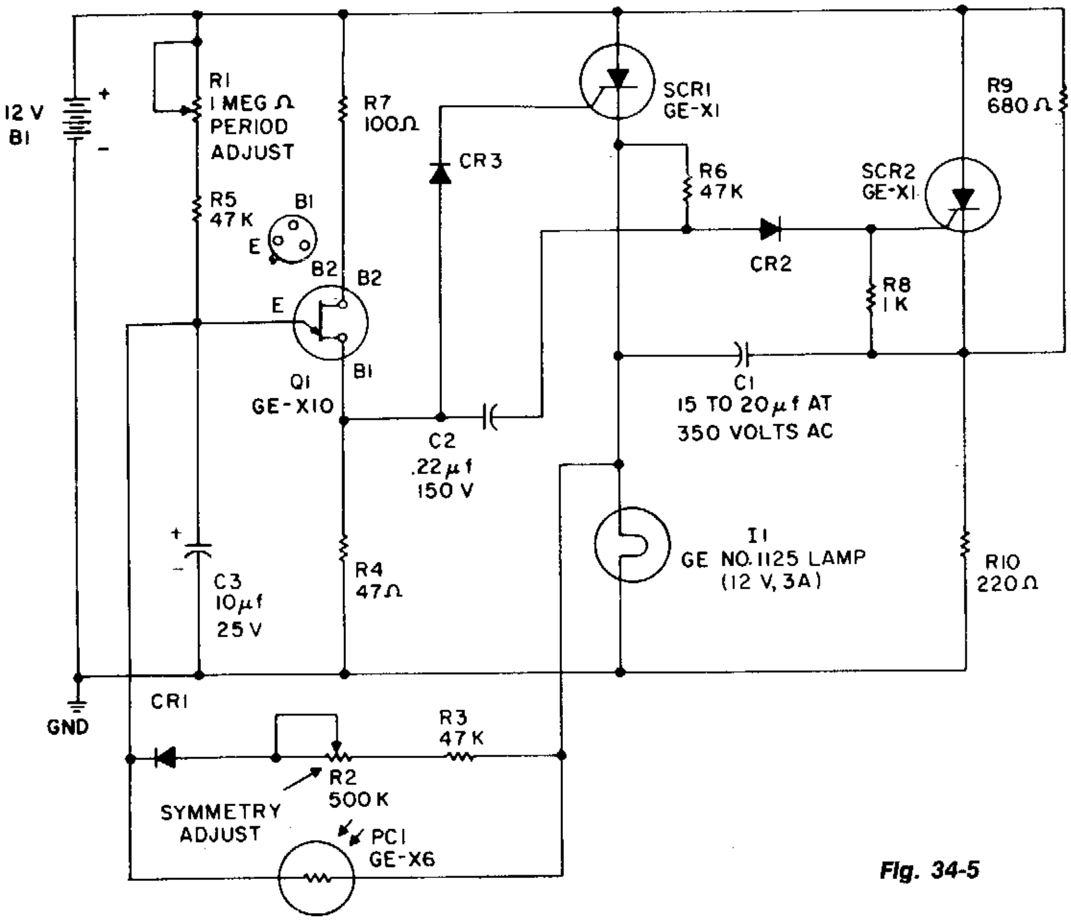


Fig. 34-5

GENERAL ELECTRIC

Circuit Notes

This flasher operates from a 12-volt car or boat battery. It offers 36 to 40-watts output, variable flash rate (up to 60 flashes per minute), independent control of both on and off cycles and photoelectric night and day control that turns the flasher on at night and shuts it off during the day for automatic operation. SCR1 and SCR2 form a basic dc flip-flop. The lamp load is the cathode leg of one SCR so that the other side of the load may be at ground (negative) potential (required in some applications). The flip-flop timing is controlled by a conventional UJT oscillator arrangement (Q1, R1, C3, etc.). Potentiometer R2 and diode CR1 provide on/off timing independence. Photoconductor PC1 locks out the UJT firing circuit during the daylight hours.

SERIES SCR LAMP FLASHER HANDLES A WIDE RANGE OF LOADS

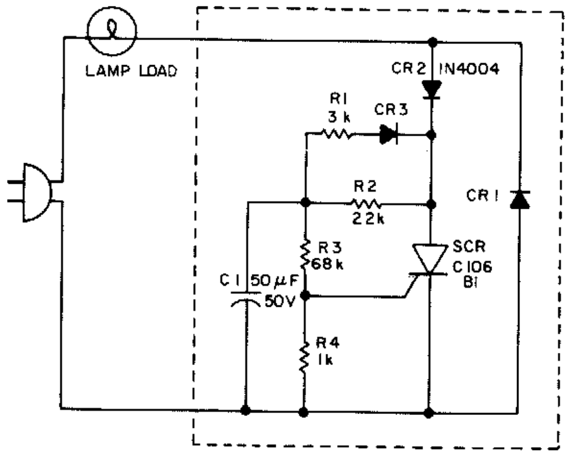


Fig. 34-6

ELECTRONIC DESIGN

Circuit Notes

Brief full-power flashes are obtained when the SCR conducts during positive half cycles of the line voltage. The SCR fires when the voltage at the divider, R3 and R4, reaches the gate-firing level. Diode D1 conducts during the reverse cycle of the SCR and provides preheating current to the lamp filaments.

SCR RELAXATION FLASHER

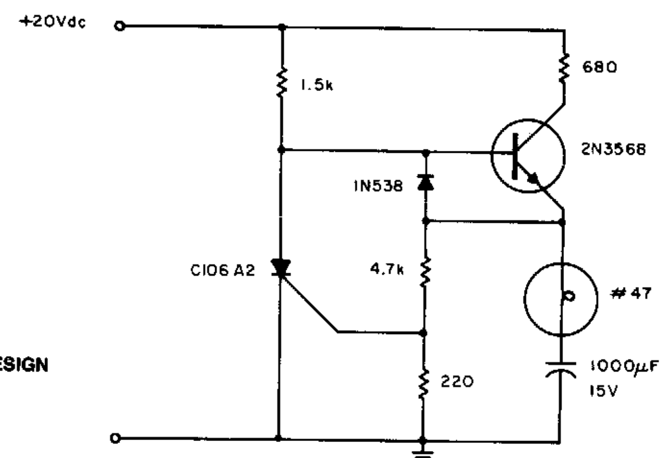


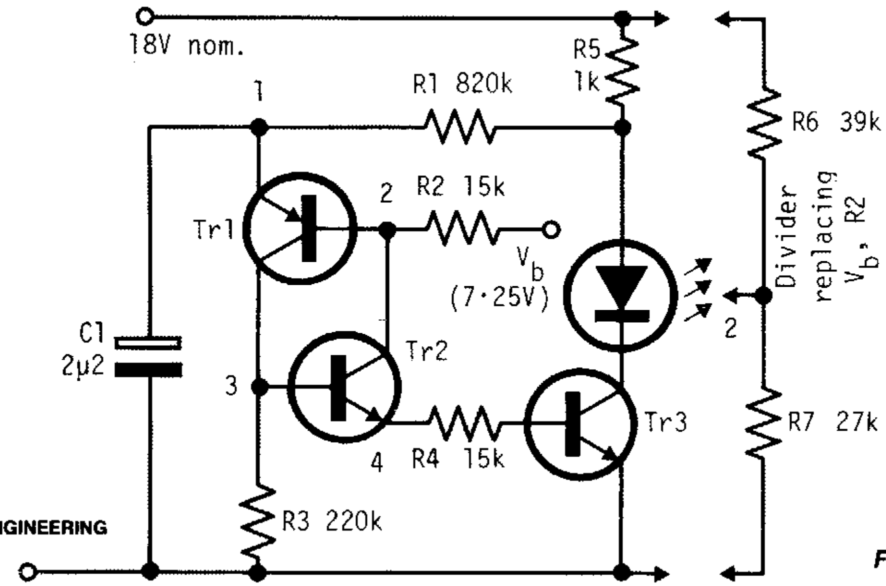
Fig. 34-7

ELECTRONIC DESIGN

Circuit Notes

Flashing occurs each time the capacitor discharges through the turned-on SCR. When the discharge current falls below the SCR holding current, the SCR turns off, and the capacitor begins charging for another cycle. The circuit will maintain a slower but good flashing capability even after considerable battery degradation.

LOW CURRENT CONSUMPTION LAMP FLASHER



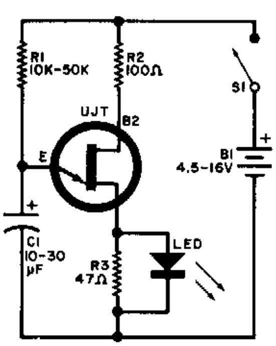
ELECTRONIC ENGINEERING

Fig. 34-8

Circuit Notes

The circuit is economical in components, and will work with virtually any transistors and is reliably self-starting. The voltage Vb can be taken from a divider, as shown at the right. If taken from a fixed source, flashing becomes slower as battery voltage falls. The lowest drive current into the base of Tr3 is about (Vb—0,6 V)/(R2 + R4). Resistor R4 limits the initial current from C1 and, as shown, R2 and R4 can be roughly equal when a divider is used for Vb. Resistor R2 equals R6R7/(R6 + R7). With the voltages shown, and with R2 = R4, the on-time is about 1.1 C1R2 and the off-time about 0.28 C1R1. Using the component values shown the period is about 0.55 sec. with a duty cycle of about 7% and a mean battery current including the Vb divider, about 1.5 mA.

LED FLASHER USES UJT



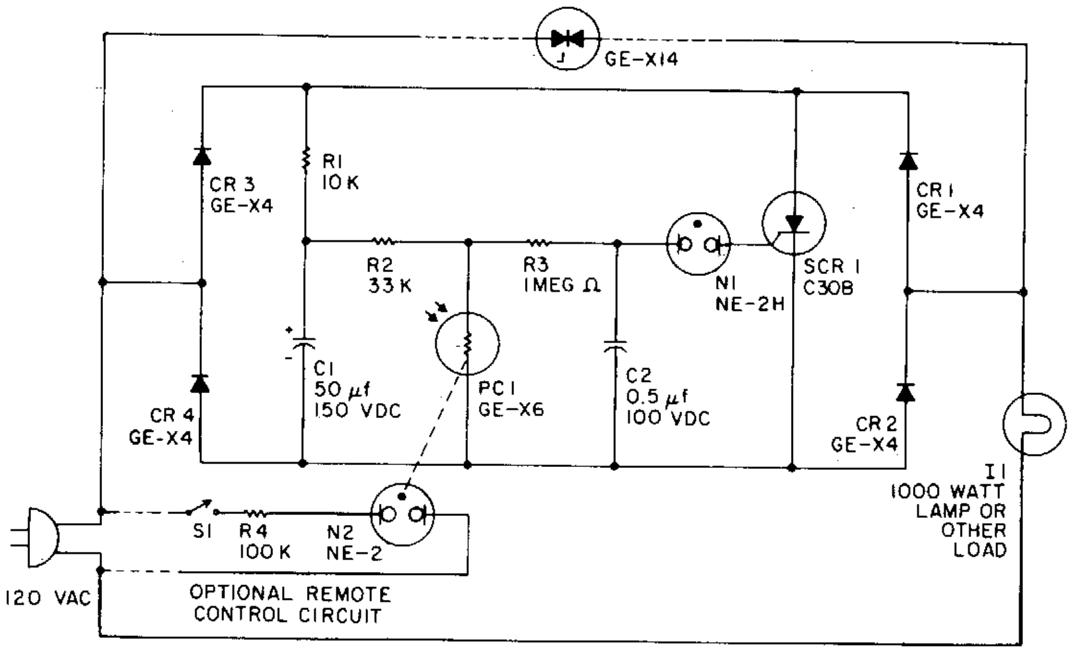
POPULAR ELECTRONICS

Fig. 34-9

Circuit Notes

A relaxation oscillator is used to flash an LED in the base circuit. C1 is charged slowly through R1 by the power source, then discharged periodically through R3 and the LED by the UJT. Flashing rate is determined by the supply voltage and by R1-C1's time constant. UJT = 2N4871

2 kW FLASHER WITH PHOTOELECTRIC CONTROL



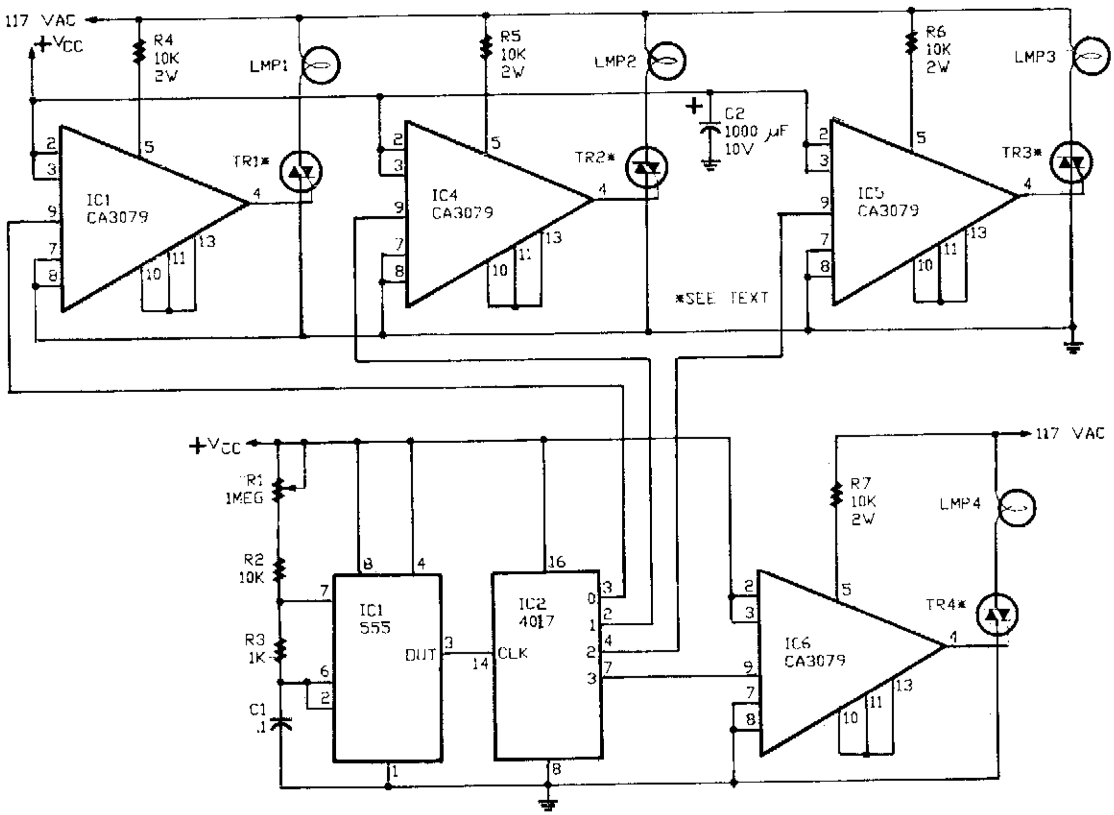
GENERAL ELECTRIC

Fig. 34-10

Circuit Notes

CR1, CR2, CR3, and CR4 form a bridge circuit with the SCR across the dc legs. With light on the photoconductor PC1, C1 charges through R1 to about 150 Vdc. The resistance of PC1 is low when illuminated, so very little voltage appears across it or C2. At about 90 volts C1 starts discharging through R1 and the SCR, but the SCR cannot turn off until C1 is almost completely discharged. When the SCR turns off during the interval line voltage is near zero, the full supply voltage again appears across the bridge, and C1 charges again to a high voltage. The voltage on C2 also starts rising until the neon lamp fires and the cycle repeats. An alternative remote control can be made by adding a second neon lamp, N2, and masking the photocell so it sees only N2. A very sensitive remote control is thus obtained that is completely isolated from the load circuit. For low-voltage remote control a flashlight lamp may be used instead of N2 and operated at about ½ its normal voltage thus giving exceptionally long life. Performance of the photoelectric control may be inverted (flash when the photoconductor is illuminated) by interchanging PC1, and R2. Sensitivity in either the normal or inverted modes can be decreased by partially masking PC1, and can be increased by increasing resistor R2 to about 470 K. To increase on time, increase C1; to increase off time, increase R3.

SEQUENTIAL FLASHER



RADIO-ELECTRONICS

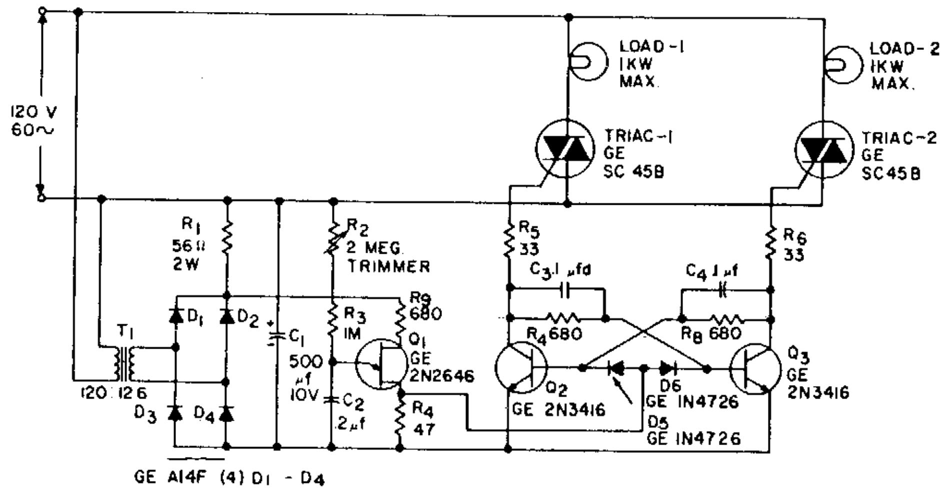
Fig. 34-11

Circuit Notes

A 555 timer, IC1, drives a 4017 CMOS decade counter. Each of the 4017's first four outputs drives a CA3079 zero-voltage switch. Pin 9 of the CA3079 is used to inhibit output from pin 4, thereby disabling the string of pulses that IC normally delivers. Those pulses occur every 8.3 ms, i.e., at a rate of 120 Hz. Each pulse has a width of 120 µs. Due to the action of the CA3079, the lamps connected to the TRI-AC's turn on and off near the zero crossing of the ac waveform. Switching at that point increases lamp life by reducing the inrush of current that would happen if the lamp were turned on near the high point of the ac waveform. In addition, switching at the zero crossing reduces Radio-Frequency Interference (RFI) considerably.

CAUTION: The CA3079's are driven directly from the 117-volt ac power line, so use care.

1 kW FLIP-FLOP FLASHER CIRCUIT



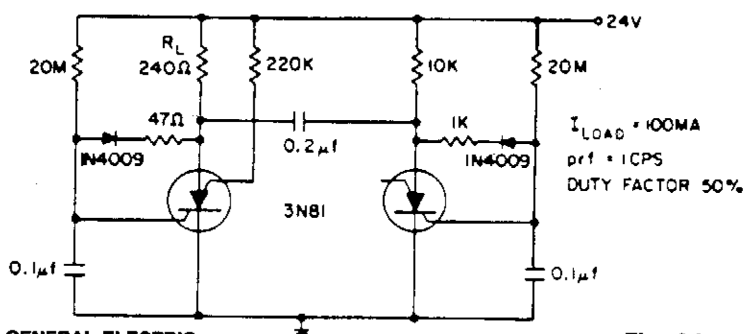
GENERAL ELECTRIC

Fig. 34-12

Circuit Notes

This is an application of the static switch circuit where the control logic is a flip-flop which is controlled by the unijunction transistor. The flashing rate can be adjusted from about 0.1 second to a 10 second cycle time.

LOW FREQUENCY OSCILLATOR-FLASHER



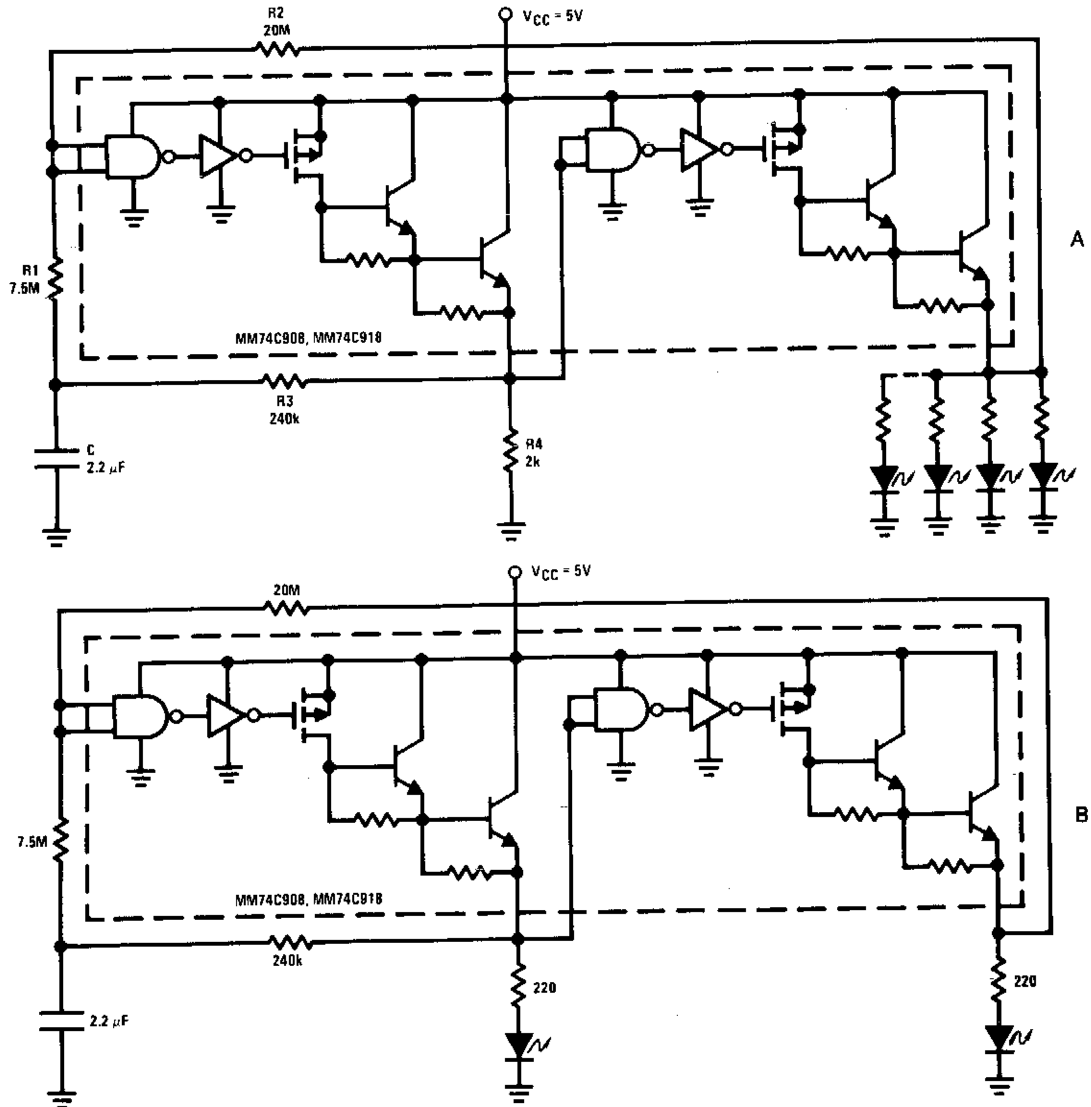
GENERAL ELECTRIC

Fig. 34-13

Circuit Notes

Electrolytic capacitors are unnecessary to generate a 1 cps frequency. As an scs triggers on, the 0.2 μF commutating capacitor turns off the other one and charges its gate capacitor to a negative potential. The gate capacitor charges towards 24 volts through 20 M retriggering its scs. Battery power is delivered to the load with 88% efficiency. The 20 M resistors can be varied to change prf or duty factor.

HIGH DRIVE OSCILLATOR/FLASHER



NATIONAL SEMICONDUCTOR CORP.

Fig. 34-14

Circuit Notes

The driver in the package is connected as a Schmitt trigger oscillator (A) where R1 and R2 are used to generate hysteresis. R3 and C are the inverting feedback timing elements and R4 is the pull-down load for the first driver. Because of its current capability, the circuit can be used to drive an array of LEDs or lamps. If resistor R4 is replaced by an LED (plus a current limiting resistor), the circuit becomes a double flasher with the 2 LEDs flashing out of phase (B).

TRANSISTORIZED FLASHERS

Flasher circuit performance as a function of temperature using limit sample transistors.			AMBIENT TEMPERATURE -40°F	77°F	212°F
CIRUIT #1	Flashes per Minute	min.	58.0	60.0	59.6
		max.	56.7	59.6	58.7
	Flash Duration in %	min.	14.5	14.6	14.3
		max.	15.1	14.9	15.3
CIRCUIT #2	Flashes per Minute	min.	58.6	60.0	59.0
		max.	55.6	58.1	56.4
	Flash Duration in %	min.	26.3	27.5	22.2
		max.	28.8	29.1	29.6
CIRCUIT #3	Flashes per Minute	min.	59.0	60.0	61.5
		max.	55.4	57.7	55.2
	Flash Duration in %	min.	45.2	46.0	48.2
		max.	45.6	46.2	47.0

Circuit Notes

Transistors Q1 and Q2 are connected as a free running multivibrator. The output, at the emitter of Q2, drives the base of the common emitter amplifier Q3, which controls the lamp. This circuit configuration permits the flash duration, the interval between flashes, and the lamp type to be varied independently. Flash duration is proportional to the product of R2C2 while the off interval is proportional to the product of R3C1. Consequently, when the flash timing must be accurately maintained, these component tolerances will have to be held to similar limits.

All three circuits described are designed for barricade warning flasher lights such as used in highway construction. They differ only in flash duration which normally is 15%, 25%, or 50% of the flash rate. Performance has been checked at ambient temperatures of −40°F, 77°F, and 212°F. A GE 5 volt, 90 milliampere type No. 1850 lamp is used.

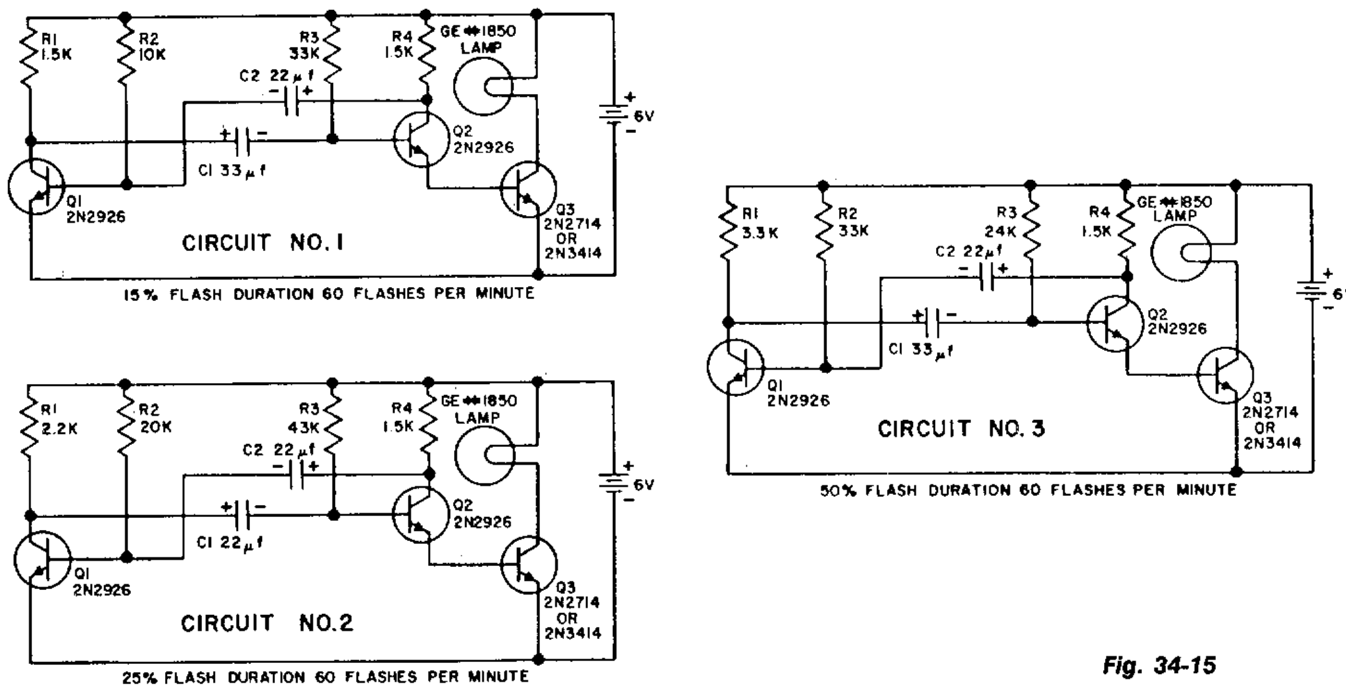


Fig. 34-15

GENERAL ELECTRIC

SEQUENTIAL AC FLASHER

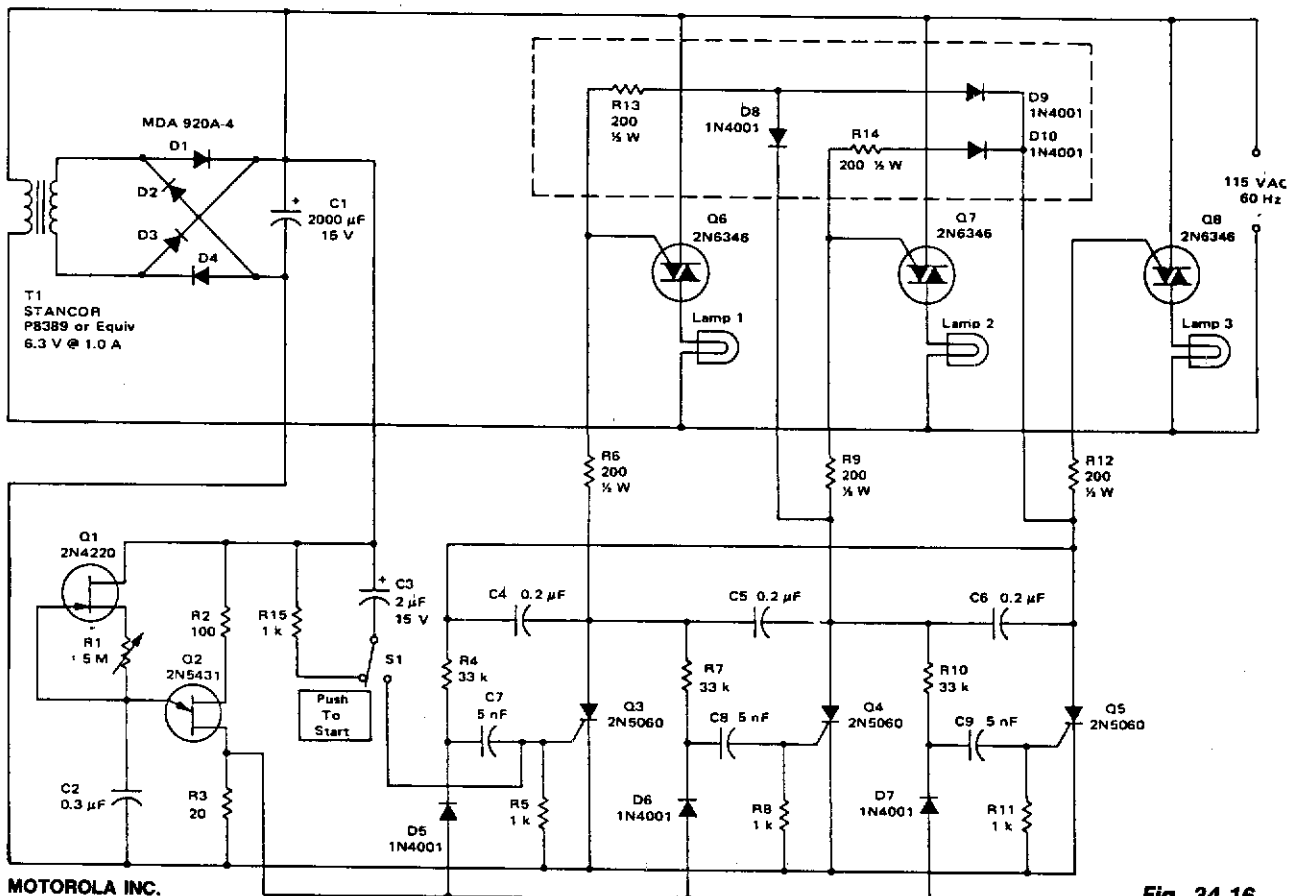
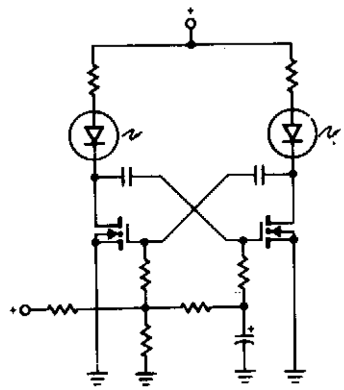


Fig. 34-16

Circuit Notes

This circuit uses a ring counter made up of Q3, Q4 and Q5. Q2 acts as an oscillator which triggers the SCRs Q3, Q4 and Q5 in sequence. When an SCR fires, a 0.2µF capacitor provides commutation, shutting off the previous stage. Triacs Q6, Q7 and Q8 control lamps 1, 2 and 3 respectively.

ASTABLE FLIP-FLOP WITH STARTER

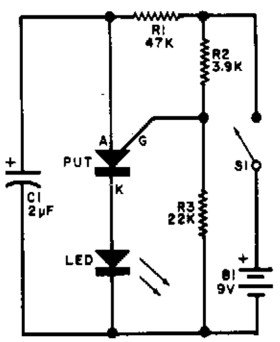


SILICONIX, INC. *Fig. 34-17*

Circuit Notes

A pair of non-zenered MOSPOWER transistors, a pair of LEDs and a simple RC circuit make an easy sequential flasher with almost unlimited sequencing time—from momentary to several seconds. The infinite input resistance of the MOSFET gate allows for very long sequencing times that are impossible when using bipolars. One precaution, though, don't wire your circuit using phenolic or printed circuit boards when looking for slow sequencing (they exhibit too much leakage!).

LED FLASHER USES PUT



POPULAR ELECTRONICS *Fig. 34-18*

Circuit Notes

This flasher circuit operates as a relaxation oscillator with C1 discharged periodically through the LED as the PUT switches on. The flashing rate is about 100/minute with the component values listed.

35

Flow Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

THERMALLY BASED ANEMOMETER (AIR FLOWMETER)

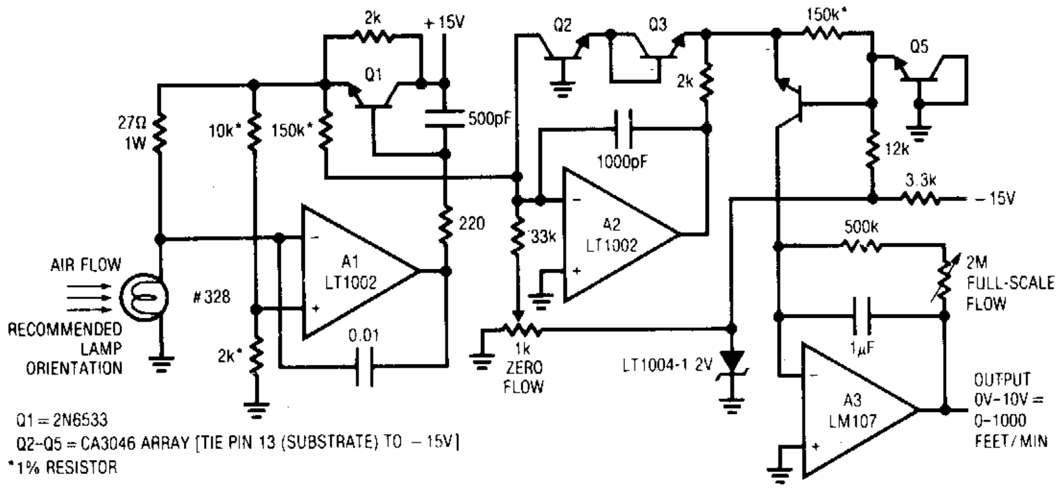


Fig. 35-1

LINEAR TECHNOLOGY CORPORATION

Circuit Notes

This design used to measure air or gas flow works by measuring the energy required to maintain a heated resistance wire at constant temperature. The positive temperature coefficient of a small lamp, in combination with its ready availability, makes it a good sensor. A type 328 lamp is modified for this circuit by removing its glass envelope. The lamp is placed in a bridge which is monitored by A1. A1's output is current amplified by Q1 and fed back to drive the bridge. When power is applied, the lamp is at a low resistance and Q1's emitter tries to come full on. As current flows through the lamp, its temperature quickly rises, forcing its resistance to increase. This action increases A1's negative input potential. Q1's emitter voltage decreases and the circuit finds a stable operating point. To keep the bridge balanced, A1 acts to force the lamp's resistance, hence its temperature, constant. The 20 k - 2 k bridge values have been chosen so that the lamp operates just below the incandescence point.

To use this circuit, place the lamp in the air flow so that its filament is at a 90° angle to the flow. Next, either shut off the air flow or shield the lamp from it and adjust the zero flow potentiometer for a circuit output of 0 V. Then, expose the lamp to air flow of 1000 feet/minute and trim the full flow potentiometer for 10 V output. Repeat these adjustments until both points are fixed. With this procedure completed, the air flowmeter is accurate within 3% over the entire 0-1000 foot/minute range.

AIR FLOW DETECTOR

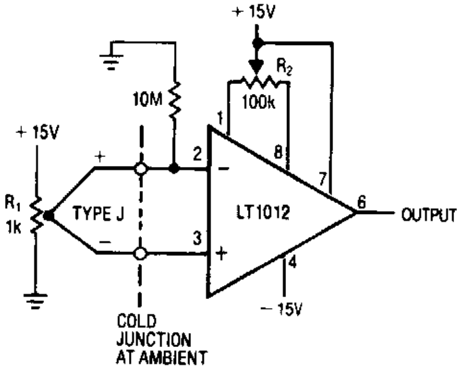


Fig. 35-2

Mount R_1 in airflow.
Adjust R_2 so output goes high when airflow stops.

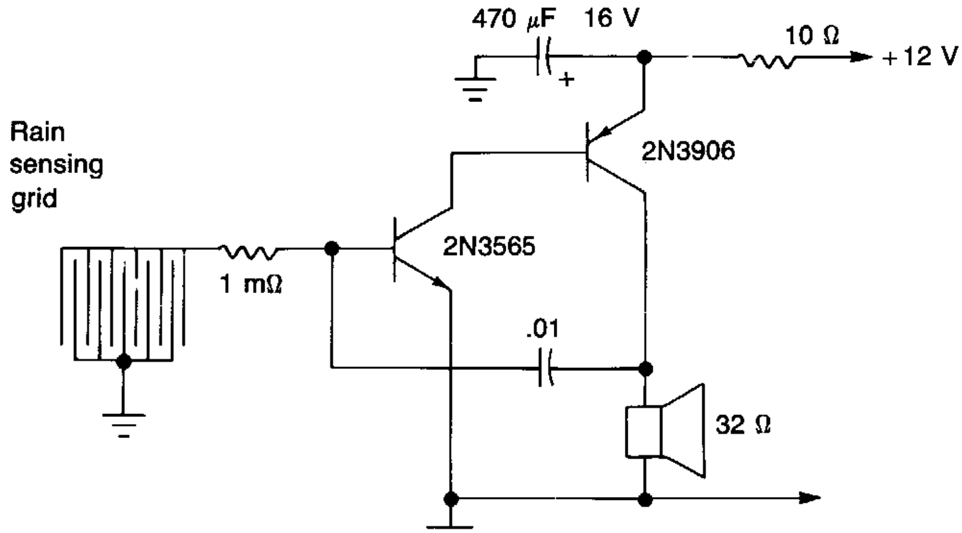
LINEAR TECHNOLOGY CORP.

36

Fluid and Moisture Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RAIN WARNING BLEEPER



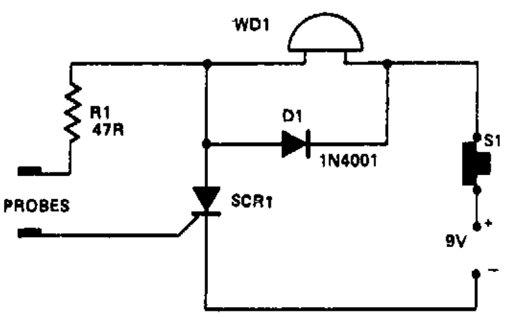
WILLIAM SHEETS

Fig. 36-1

Circuit Notes

One small spot of rain on the sense pad of this bleeper will start this audio warning. It can also be operated by rising water. The circuit has two transistors, with feedback via capacitor C1, but Trl cannot operate as long as the moisture sense pad is dry. When the pad conducts, Trl and Tr2 form an audio oscillatory circuit, the pitch depends somewhat on the resistance.

WATER-LEVEL INDICATOR



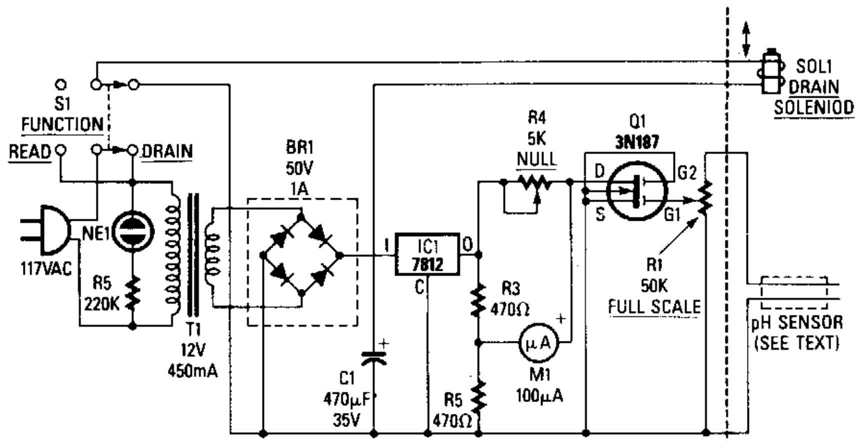
ELECTRONICS TODAY INTERNATIONAL

Fig. 36-2

Circuit Notes

In this a warning device WD1 is in series with SCR1. When the liquid level causes a conductive path between the probes, the SCR conducts sounding WD1. The warning device may be a Sonalert (TM), a lamp or a buzzer. D1 acts as a transient suppressor. Press S1 to reset the circuit.

ACID RAIN MONITOR



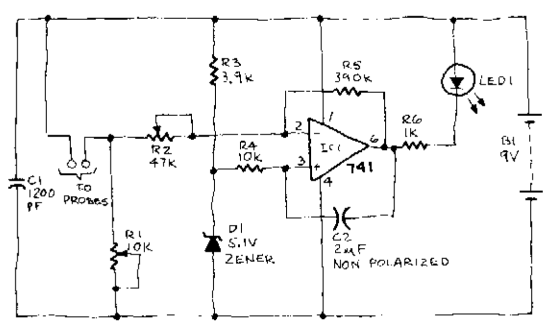
RADIO-ELECTRONICS

Fig. 36-3

Circuit Notes

A bridge rectifier and 12-volt regulator powers the MOSFET sensing circuit. The unregulated output of the bridge rectifier operates the drain solenoid via switch S1. The sensor itself is built from two electrodes, one made of copper, the other of lead. In combination with the liquid trapped by the sensor, they form a miniature lead-acid cell whose output is amplified by MOSFET Q1. The maximum output produced by our prototype cell was about 50 μA. MOSFET Q1 serves as the fourth leg of a Wheatstone bridge. When acidity causes the sensor to generate a voltage, Q1 turns on slightly, so its drain-to-source resistance decreases. That resistance variation causes an imbalance in the bridge, and that imbalance is indicated by meter M1.

PLANT-WATER MONITOR



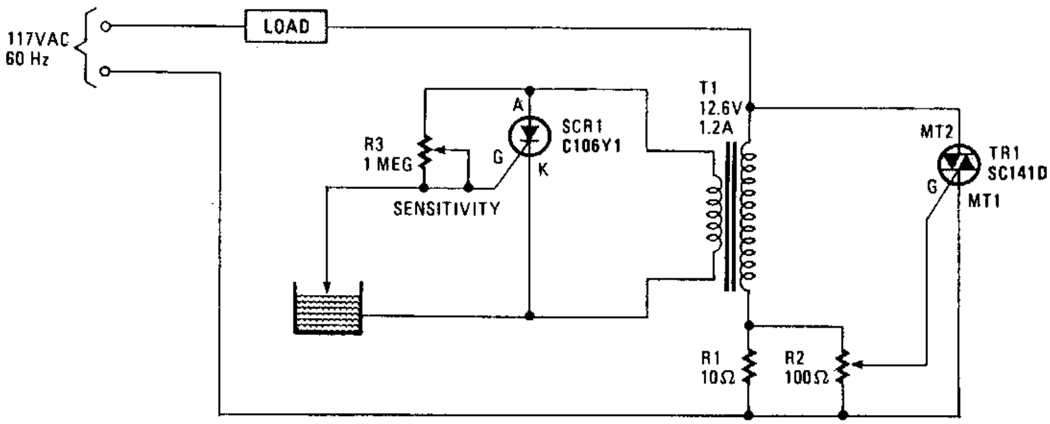
RADIO-ELECTRONICS

Fig. 36-4

Circuit Notes

When the soil is moist, the LED glows. If the moisture falls below a certain predetermined level, the LED begins to flash. If there is still less moisture, the LED turns off. To calibrate, connect the battery and insert the probe into a container of dry soil. Set R1 to its maximum value then reduce that resistance until the LED begins to flash. The range over which the LED flashes before going out is adjusted using R2.

WATER-LEVEL SENSING AND CONTROL

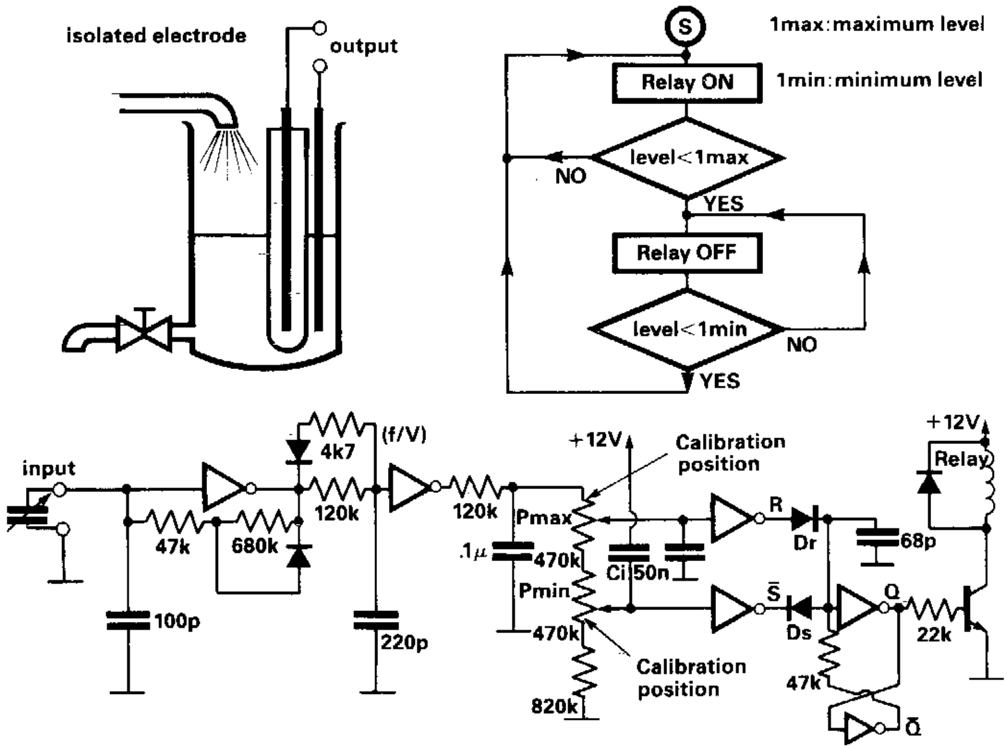


HANDS-ON ELECTRONICS *Fig. 36-5*

Circuit Notes

The operation of the circuit is based on the difference in the primary impedance of a transformer when its secondary is loaded and when it is open-circuit. The impedance of the primary of T1 and resistor R1 are in series with the load. The triac's gate-control voltage is developed across parallel resistors R1 and R2. When the water level is low, the probe is out of the water and SCR1 is triggered on. It conducts and imposes a heavy load on transformer T1's secondary winding. That load is reflected back into the primary, gating triac TR1 on, which energizes the load. If the load is an electric value in the water-supply line, it will open and remain open until the water rises and touches the probe, which shorts SCR1's gate and cathode, thereby turning off the SCR1, which effectively open-circuits the secondary. The open-circuit condition—when reflected back to the primary winding—removes the triac's trigger signal, thereby turning the water off. The load may range from a water valve, a relay controlling a pump supplying water for irrigation, or a solenoid valve controlling the water level in a garden lily pond.

SINGLE CHIP PUMP CONTROLLER



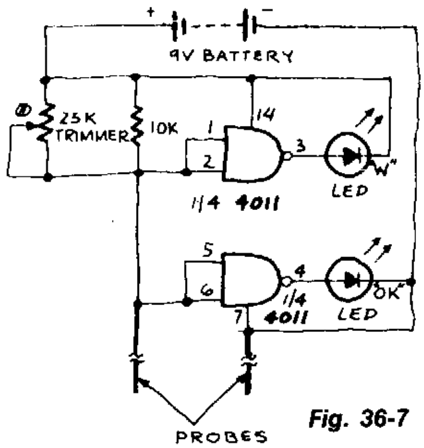
ELECTRONIC ENGINEERING

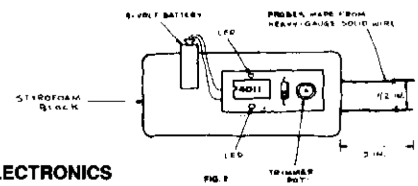
Fig. 36-6

Circuit Notes

This circuit controls the level of a tank using a bang-bang controlled electrical pump. The actual level of liquid is measured by a capacitive level-meter. The first inverter performs as a capacitance to frequency converter. It is a Schmitt oscillator and its frequency output decreases as the capacitance increases. The second inverter is a monostable which performs as a frequency to voltage converter (f/V). Its output is applied to the maximum and minimum level comparator inputs. Maximum and minimum liquid levels may be set by the potentiometers. The maximum level (1 max) may be preset between the limits: 65 pF less than C (1 max) less than 120 pF. The minimum level is presetable and the limits are: 0 less than C (1 min) less than 25 pF.

PLANT WATER GAUGE





RADIO-ELECTRONICS

Fig. 36-7

Circuit Notes

To calibrate the gauge, connect the battery and press the probes gently into a pot containing a plant that is just on the verge of needing water (stick it in so that only an inch of the probe is left visible at the top). Turn the potentiometer until the "OK" LED lights and then turn it back to the point where that LED goes out and the "W", or "Water", LED just comes on. The device should now be properly adjusted.

LIQUID FLOWMETER

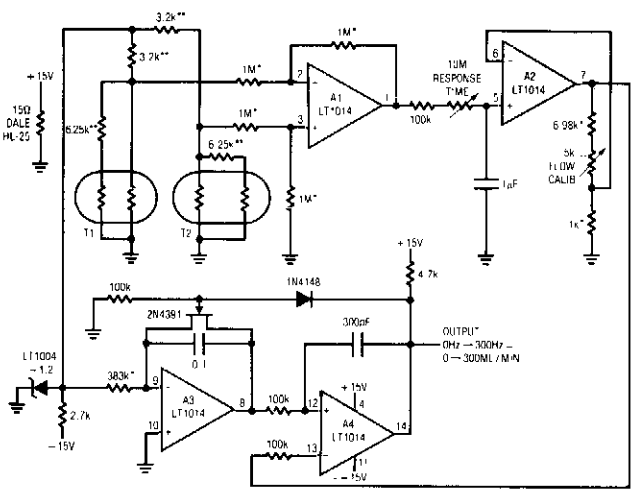
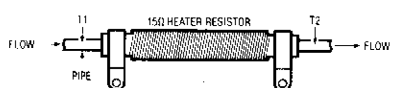


Fig. 36-8



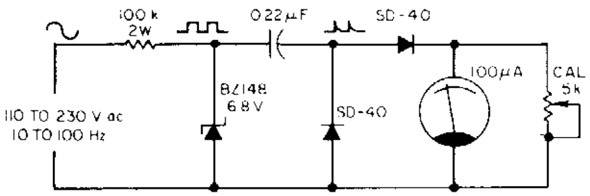
*1% FILM RESISTOR
**SUPPLIED WITH YSI THERMISTOR NETWORK
T1, T2 YSI THERMISTOR NETWORK = #44201
FLOW IN PIPE IS INVERSELY PROPORTIONAL TO RESISTANCE OF T1–T2 TEMPERATURE DIFFERENCE.
A1–A2 PROVIDE GAIN. A3, A4 PROVIDE LINEARIZED FREQUENCY OUTPUT.

LINEAR TECHNOLOGY CORP.

37
Frequency Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

POWER-FREQUENCY METER



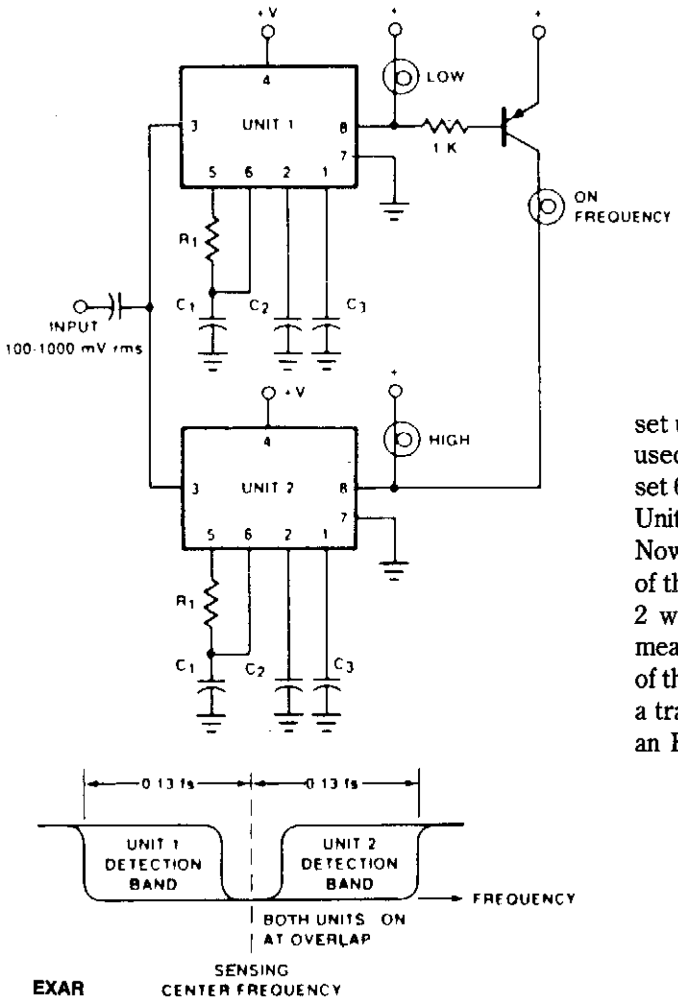
ELECTRONIC DESIGN

Fig. 37-1

Circuit Notes

The meter uses a zener diode to form square waves from input sine waves. After calibration with the 5 k ohm potentiometer, the 100 μA meter reads directly in hertz.

LOW COST FREQUENCY INDICATOR



EXAR

Fig. 37-2

Circuit Notes

The circuit shows how two tone decoders set up with overlapping detection bands can be used for a go/no go frequency meter. Unit 1 is set 6% above the desired sensing frequency and Unit 2 is set 6% below the desired frequency. Now, if the incoming frequency is within 13% of the desired frequency, either Unit 1 or Unit 2 will give an output. If both units are on, it means that the incoming frequency is within 1% of the desired frequency. Three light bulbs and a transistor allow low cost read-out. The IC is an EXAR 567.

38

Frequency Multiplier and Divider Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

NONSELECTIVE FREQUENCY TRIPLER USES TRANSISTOR SATURATION CHARACTERISTICS

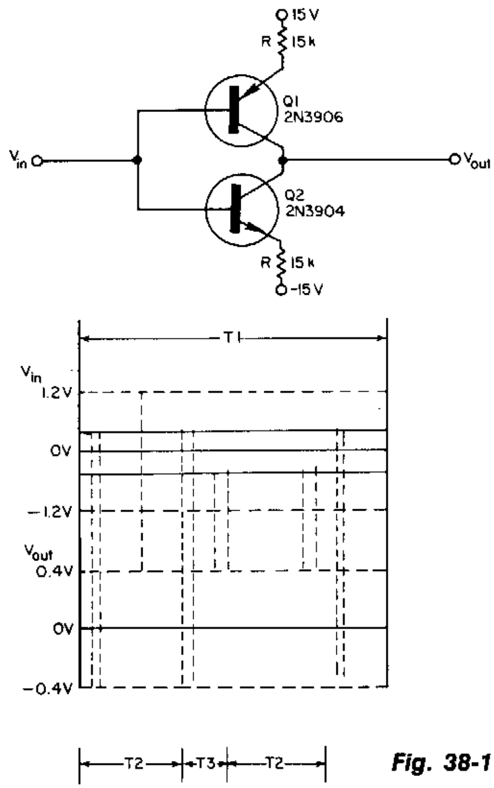


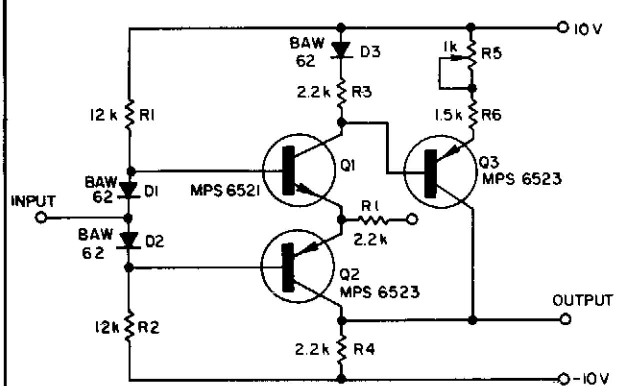
Fig. 38-1

ELECTRONIC DESIGN

Circuit Notes

The turn-on and turn-off characteristics of two complementary transistors can be combined to attain nonselective frequency tripling. The resulting circuit handles any periodic waveform with nonvertical sides. Each input signal peak produces three output signal peaks. The additional peaks occur where the input signal causes saturation of one of the two transistors. The circuit operates over a frequency range from dc to the upper limits of the complementary transistor pair. About the only disadvantage of the circuit is the lack of symmetry of the output signal peaks.

FREQUENCY DOUBLER WORKS TO 1 MHz



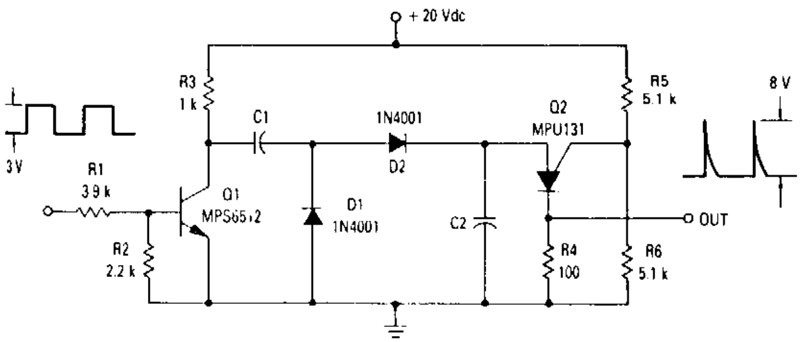
ELECTRONIC DESIGN

Fig. 38-2

Circuit Notes

Adding components Q3, D3, and resistors R3 through R6 to a conventional complementary symmetry class AB buffer can double the frequency of an input sine wave.

LOW FREQUENCY DIVIDER



TABLE

C_1	C_2	Division
0.01 μF	0.01 μF	2
0.01 μF	0.02 μF	3
0.01 μF	0.03 μF	4
0.01 μF	0.04 μF	5
0.01 μF	0.05 μF	6
0.01 μF	0.06 μF	7
0.01 μF	0.07 μF	8
0.01 μF	0.08 μF	9
0.01 μF	0.09 μF	10
0.01 μF	0.1 μF	11

MOTOROLA ***Fig. 38-3***

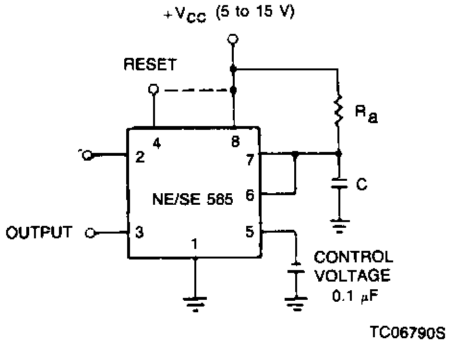
Circuit Notes

The ratio of capacitors C1 and C2 determines division. With a positive pulse applied to the base of Q1, assume that C1 = C2 and that C1 and C2 are discharged. When Q1 turns off, both C1 and C2 charge to 10 volts each through R3. On the next pulse to the base of Q1, C1 is again discharged but C2 remains charged to 10 volts. As Q1 turns off this time, C1 and C2 again charge. This time C2 charges to the peak point firing voltage of the PUT causing it to fire. This discharges capacitor C2 and allows capacitor C1 to charge to the line voltage. As soon as C2 discharges and C1 charges, the PUT turns off. The next cycle begins with another positive pulse on the base of Q1 which again discharges C1. The input and output frequency can be approximated by the equation

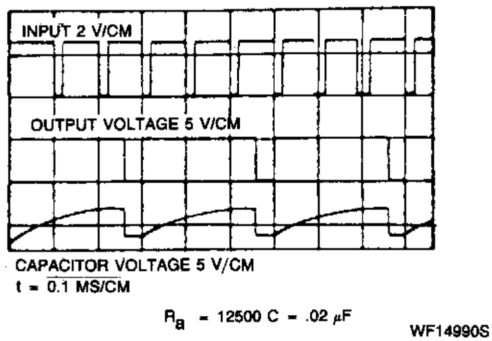
$$f_{in} = \frac{(C1 = C2)}{C1} f_{out}$$

For a 10 kHz input frequency with an amplitude of 3 volts, the table shows the values for C1 and C2 needed to divide by 2 to 11.

FREQUENCY DIVIDER



Schematic Diagram



Expected Waveforms

SIGNETICS

Fig. 38-4

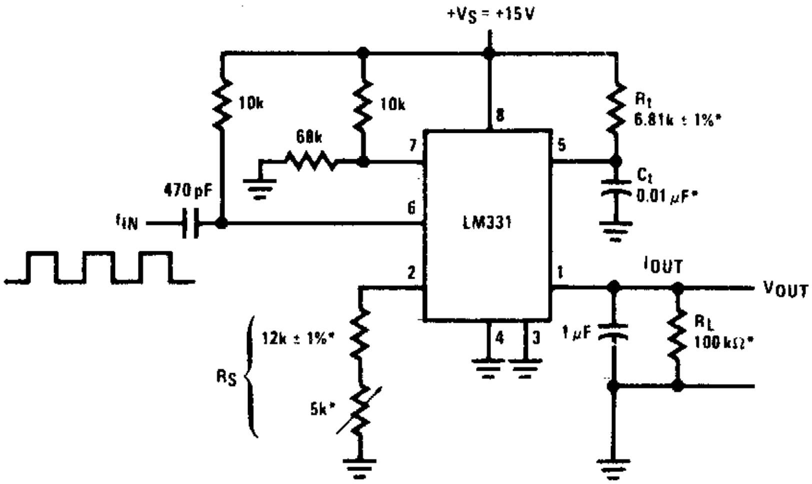
Circuit Notes

If the input frequency is known, the timer can easily be used as a frequency divider by adjusting the length of the timeing cycle. Figure shows the waveforms of the timer when used as a divide-by-three circuit. This application makes use of the fact that this circuit cannot be retriggered during the timing cycle.

39

Frequency-to-Voltage Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.



$$V_{OUT} = f_{IN} \times 2.09V \times \frac{R_L}{R_S} \times (R_tC_t)$$

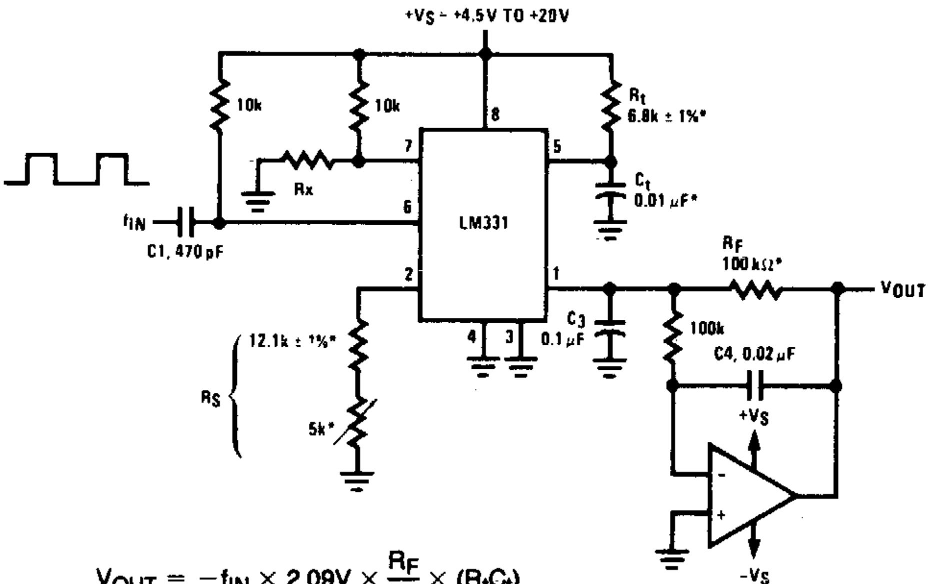
*Use stable components with low temperature coefficients.

Simple Frequency-to-Voltage Converter, 10 kHz Full-Scale, ±0.06% Non-Linearity

NATIONAL SEMICONDUCTOR CORP. *Fig. 39-1*

Circuit Notes

In these applications, a pulse input at f_{IN} is differentiated by a C-R network and the negative-going edge at pin 6 causes the input comparator to trigger the timer circuit. Just as with a V-to-F converter, the average current flowing out of pin 1 is $I_{AVERAGE} = i \times (1.1\ R_1C_1) \times f$. In this simple circuit, this current is filtered in the network R_L = 100 k ohm and 1 μF. The ripple will be less than 10 mV peak, but the response will be slow, with a 0.1 second time constant, and settling of 0.7 second to 0.1%



$$V_{OUT} = -f_{IN} \times 2.09V \times \frac{R_F}{R_S} \times (R_tC_t)$$

TL/H/5680–8

$$\text{SELECT } Rx = \frac{(V_S - 2V)}{0.2\text{ mA}}$$

***Use stable components with low temperature coefficients.**

Precision Frequency-to-Voltage Converter, 10 kHz Full-Scale with 2-Pole Filter, ±0.01% Non-Linearity Maximum

accuracy. In the precision circuit, an operational amplifier provides a buffered output and also acts as a 2-pole filter. The ripple will be less than 5 mV peak for all frequencies above 1 kHz, and the response time will be much quicker than in Part 1. However, for input frequencies below 200 Hz, this circuit will have worse ripple than the figure. The engineering of the filter time-constants to get adequate response and small enough ripple simply requires a study of the compromises to be made. Inherently, V-to-F converter response can be fast, but F-to-V response cannot.

40
Function Generator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

QUAD OP AMP GENERATES FOUR DIFFERENT SYNCHRONIZED WAVEFORMS SIMULTANEOUSLY

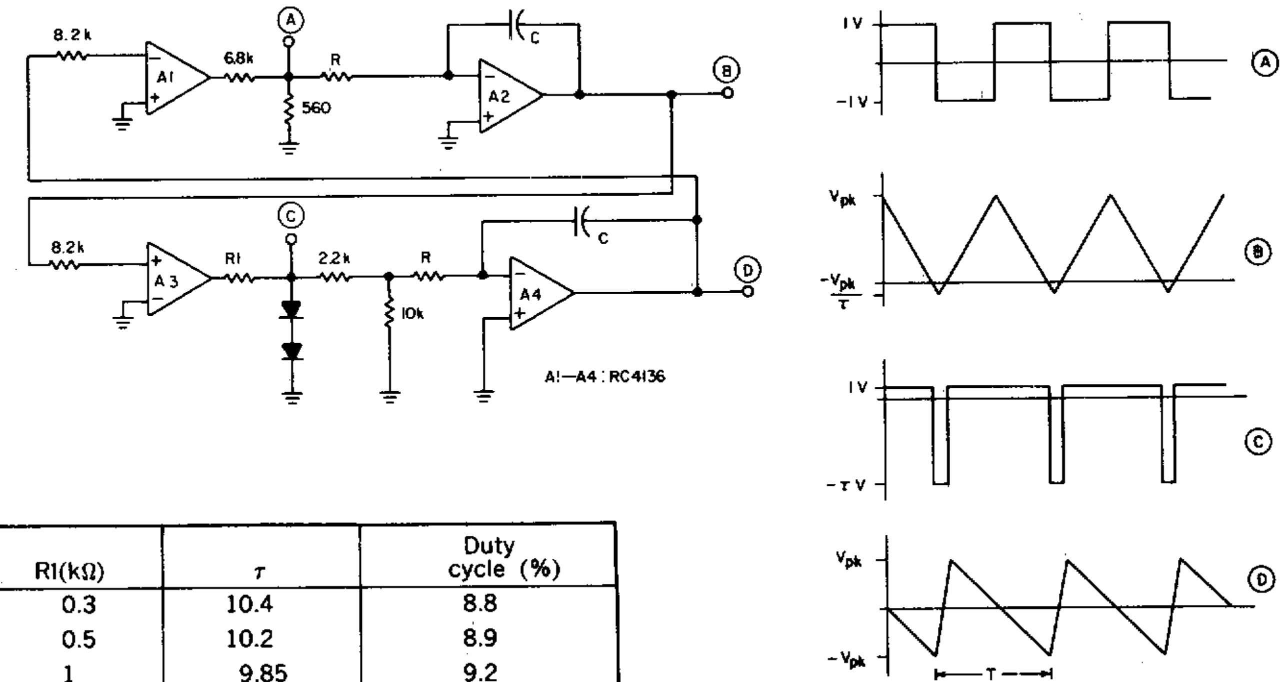


Fig. 40-1

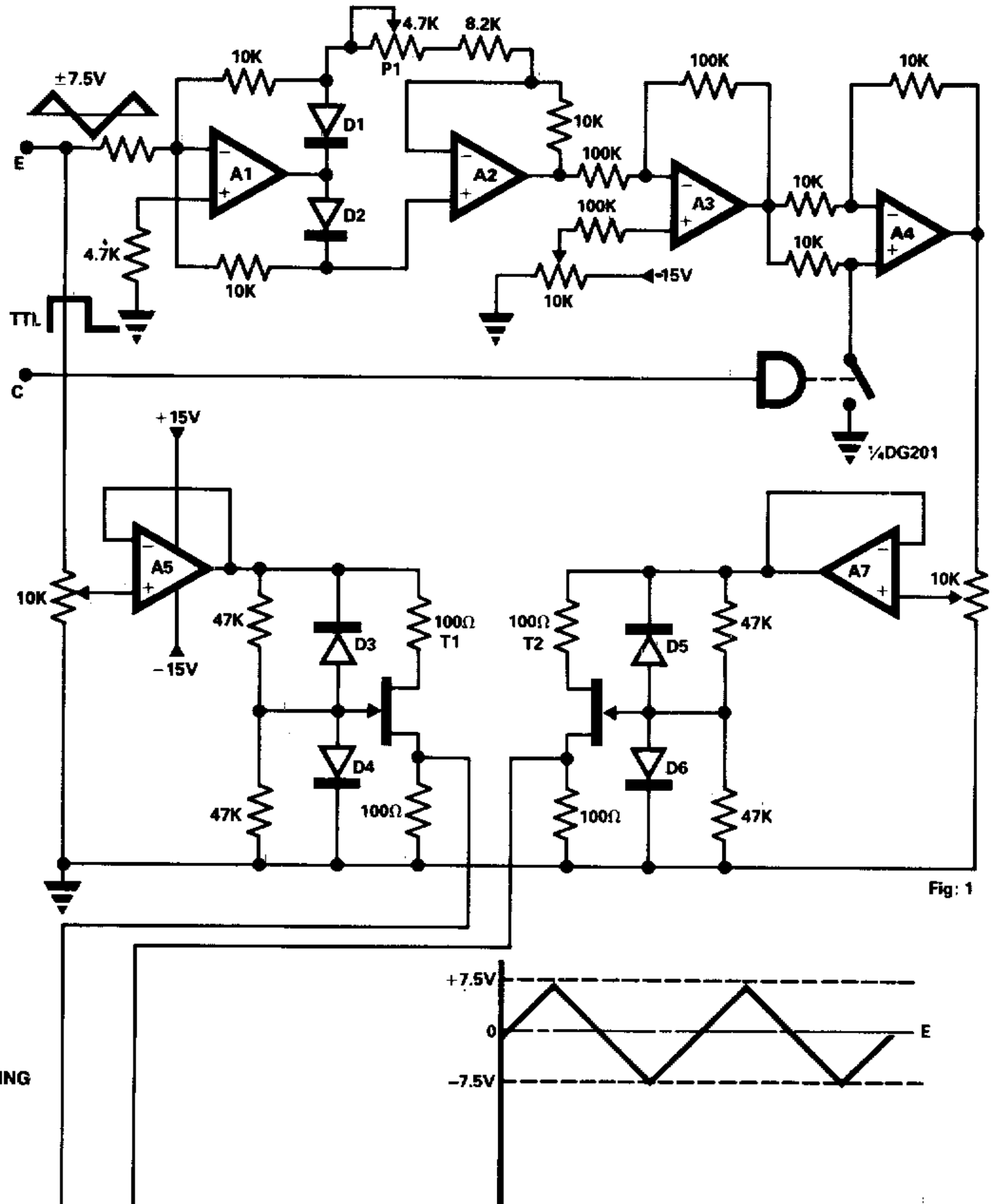
R1(kΩ)	τ	Duty cycle (%)
0.3	10.4	8.8
0.5	10.2	8.9
1	9.85	9.2
4	8.0	11.1
10	5.86	14.6
20	4.04	19.84
50	2.09	32.4
100	1.16	46.30
117.8	1	50

ELECTRONIC DESIGN

Circuit Notes

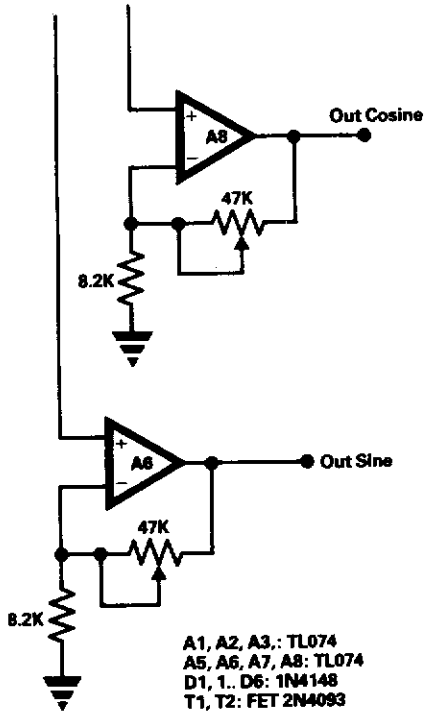
A quad op amp can simultaneously generate four synchronized waveforms. The two comparators (A1 and A3) produce square and pulse waves, while the two integrators (A2 and A4) give triangular and sawtooth waves. Resistor R1 sets the duty cycle and the frequency, along with resistors R and capacitors C.

A SINE/COSINE GENERATOR FOR 0.1 - 10 kHz



ELECTRONIC ENGINEERING

Fig. 40-2



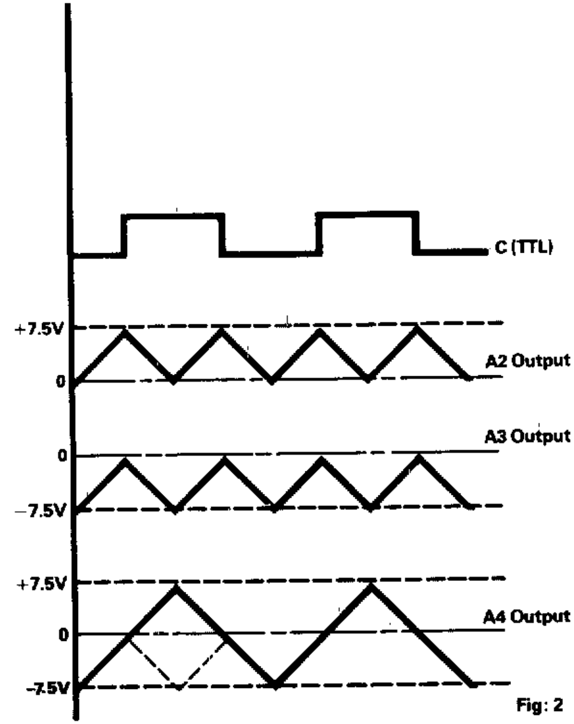


Fig: 2

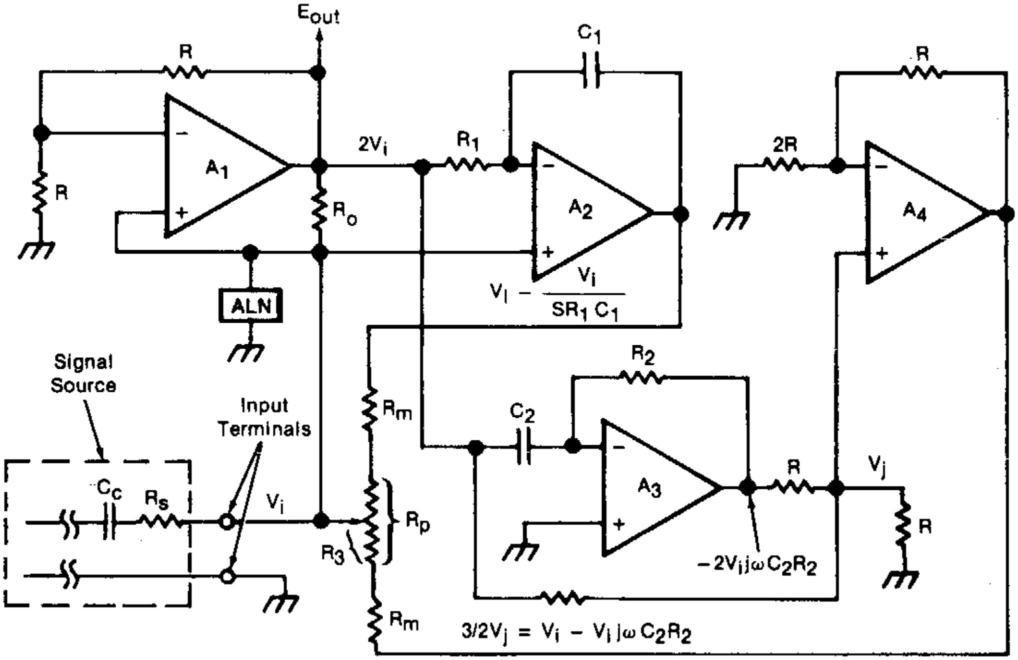
Circuit Notes

The scheme presented delivers waveforms from any function generator producing a triangular output and a synchronized TTL square wave. A1 and A2 act as a two-phase current rectifier by inverting the negative voltage appearing at the input of A1.

Positive input: Both A1 and A2 work as unity gain followers, D1 and D2 being in the off-state.

Negative input: A1 has a $-2/3$ gain (D1 off and D2 on), A2 has a $+3/2$ gain and the total voltage transfer is -1 between output and input. P1 allows a fine trimming of the -1 gain for the negative input signals. A3 adds a continuous voltage to the rectified positive signal in order to attack A4 which acts as a $\pm$ multiplier commanded by the TTL input through the analog switch. The signal polarity is reconstructed and the output of A4 delivers a triangular waveform shifted by 90° with respect to the input signal, Fig. 2. The original and the shifted voltages are fed into the triangle to sine converters through A5 and A7 working as impedance converters. Over the frequency dynamic ranges from 0.1 Hz to 10 kHz, the phase shift is constant and the distortion on the sine voltage is less than 1%.

OSCILLATOR OR AMPLIFIER WITH WIDE FREQUENCY RANGE



NASA

Fig. 40-3

Circuit Notes

An oscillator/amplifier is resistively tunable over a wide frequency range. Feedback circuits containing operational amplifiers, resistors, and capacitors synthesize the electrical effects of an inductance and capacitance in parallel between the input terminals. The synthetic inductance and capacitance, and, therefore, the resonant frequency of the input admittance, are adjusted by changing a potentiometer setting. The input signal is introduced in parallel to the noninverting input terminals of operational amplifiers A_1 and A_2 and to the potentiometer cursor. The voltages produced by the feedback circuits in response to input voltage V_1 are indicated at the various circuit nodes.

LINEAR TRIANGLE/SQUARE WAVE VCO

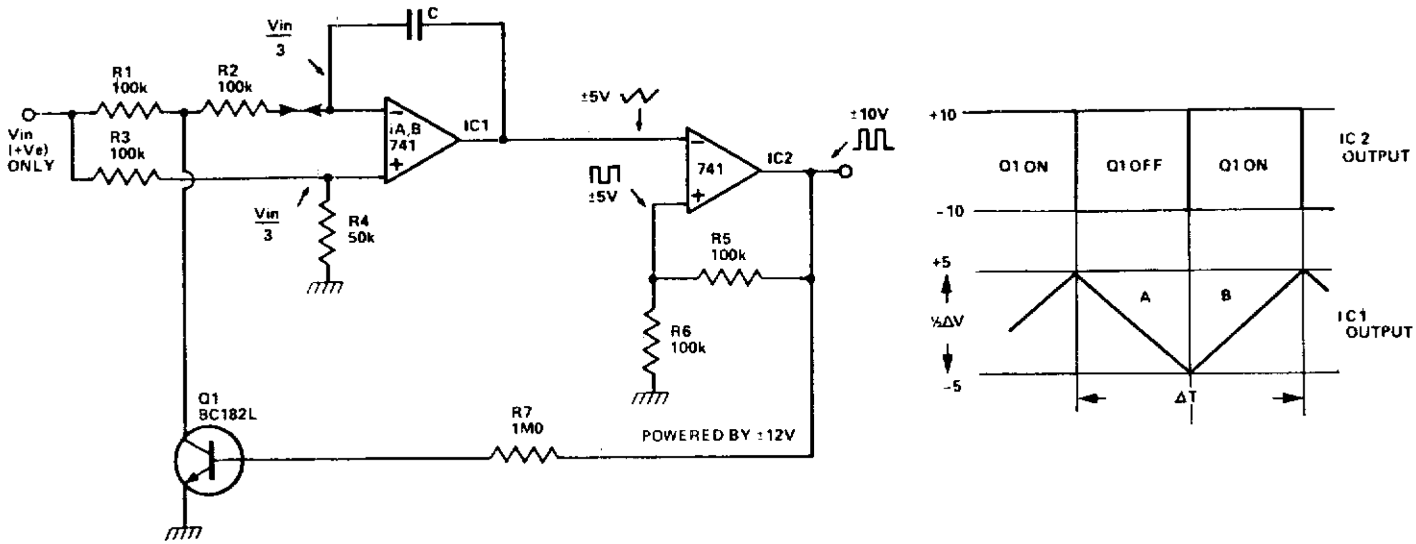


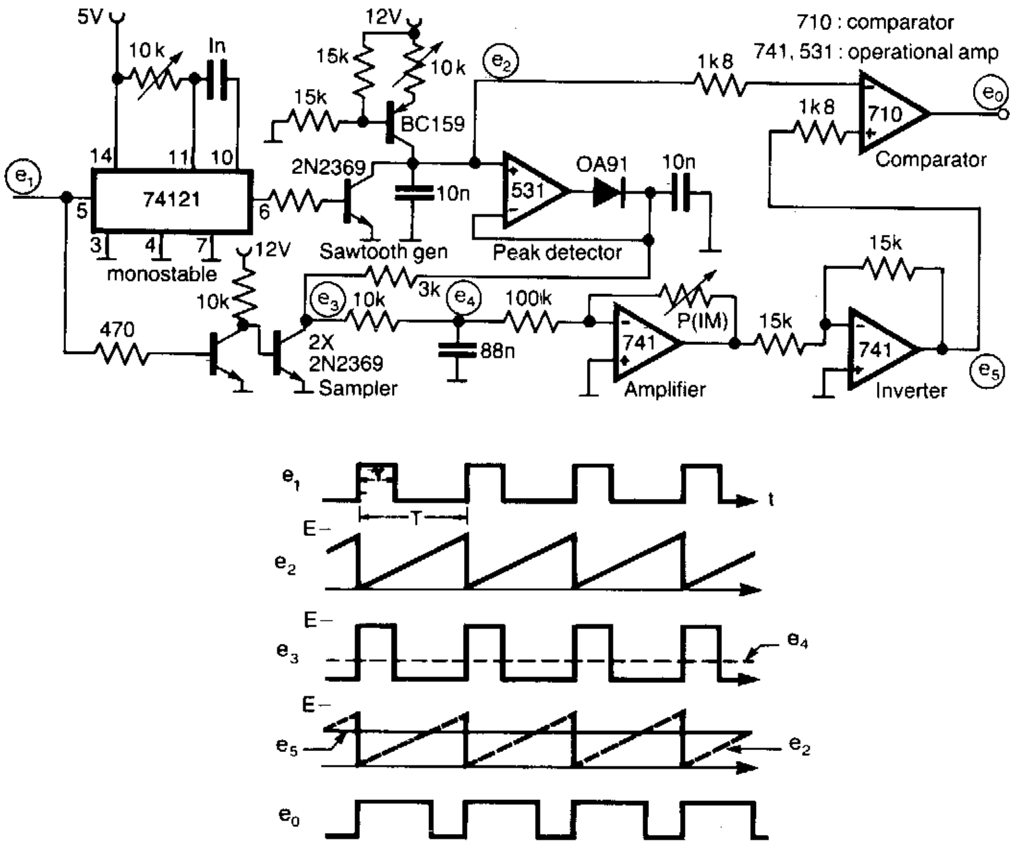
Fig. 40-4

ELECTRONICS TODAY INTERNATIONAL

Circuit Notes

The VCO has two buffered outputs; a triangle wave and a square wave. Frequency is dependent on the output voltage swing of the Schmitt trigger, IC2. Superior performance can be obtained by replacing Q1 with a switching FET. Fast FET op amps will improve high frequency performance.

CIRCUIT FOR MULTIPLYING PULSE WIDTHS



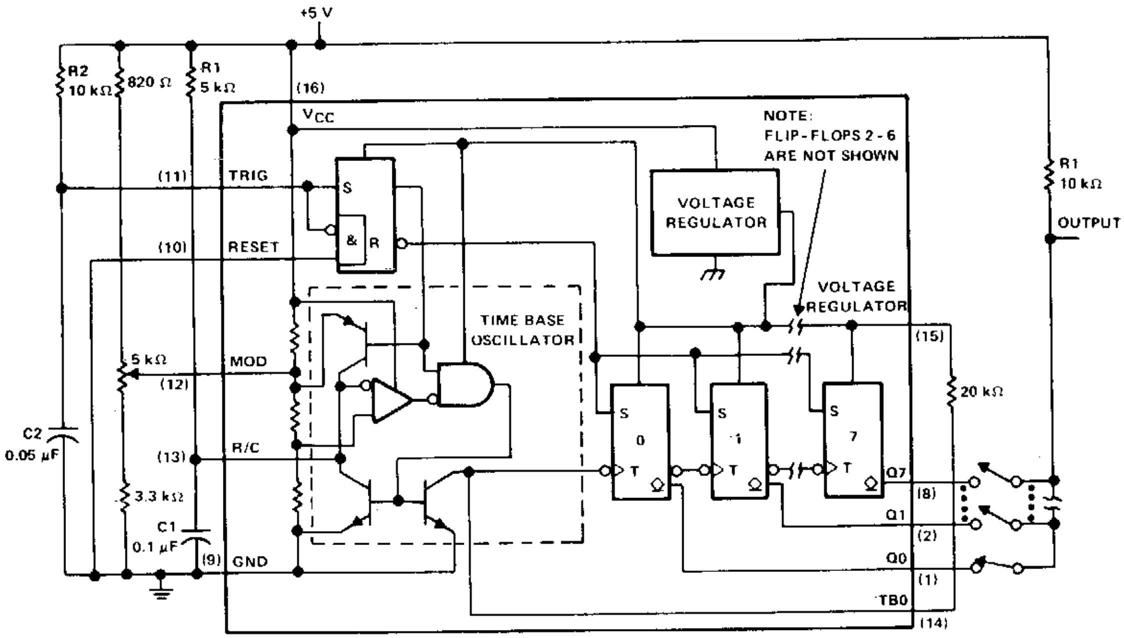
ELECTRONIC ENGINEERING

Fig. 40-5

Circuit Notes

A circuit for multiplying the width of incoming pulses by a factor greater or less than unity is simple to build and has the feature that the multiplying factor can be selected by adjusting one potentiometer only. The multiplying factor is determined by setting the potentiometer P in the feedback of a 741 amplifier. The input pulses e_1 of width τ and repetition period T is used to trigger a sawtooth generator at its rising edges to produce the waveform e_2 having a peak value of (E) volt. This peak value is then sampled by the input pulses to generate the pulse train e_3 having an average value of $e_4(= E\ E\tau/T)$ which is proportional to τ and independent on T. The dc voltage e_4 is amplified by a factor k and compared with sawtooth waveform e_2 giving output pulses of duration k τ. The circuit is capable of operating over the frequency range 10 kHz - 100 kHz.

PROGRAMMABLE VOLTAGE CONTROLLED FREQUENCY SYNTHESIZER



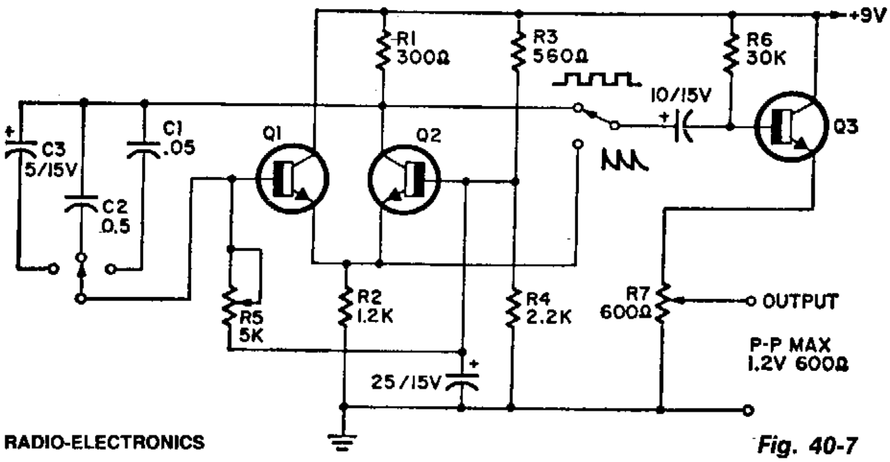
TEXAS INSTRUMENTS

Fig. 40-6

Circuit Notes

The μA2240 consists of four basic circuit elements: (1) a time-base oscillator, (2) an eight-bit counter, (3) a control flip-flop, and (4) a voltage regulator. The basic frequency of the time-base oscillator (TBO) is set by the external time constant determined by the values of R1 and C1 (1?R1C1 = 2 kHz). The open-collector output of the TBO is connected to the regulator output via a 20 k ohm pull-up resistor, and drives the input to the eight-bit counter. At power-up, a positive trigger pulse is detected across C2 which starts the TBO and sets all counter outputs to a low state. Once the μA2240 is initially triggered, any further trigger inputs are ignored until it is reset. In this astable operation, the μA2240 will free-run from the time it is triggered until it receives an external reset signal. Up to 255 discrete frequencies can be synthesized by connecting different counter outputs.

EMITTER-COUPLED RC OSCILLATOR



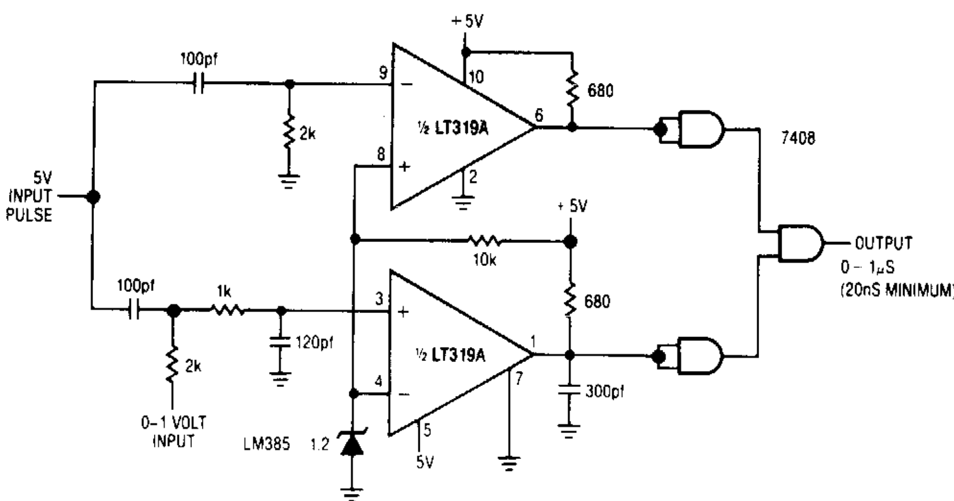
RADIO-ELECTRONICS

Fig. 40-7

Circuit Notes

The circuit covers 15 Hz-30 kHz and is useful as a function generator. The 2N2926 or equivalent transistors can be used.

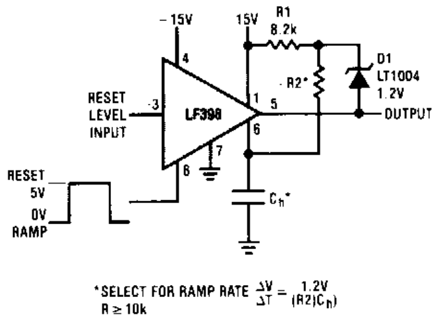
VOLTAGE CONTROLLED HIGH SPEED ONE SHOT



LINEAR TECHNOLOGY CORP.

Fig. 40-8

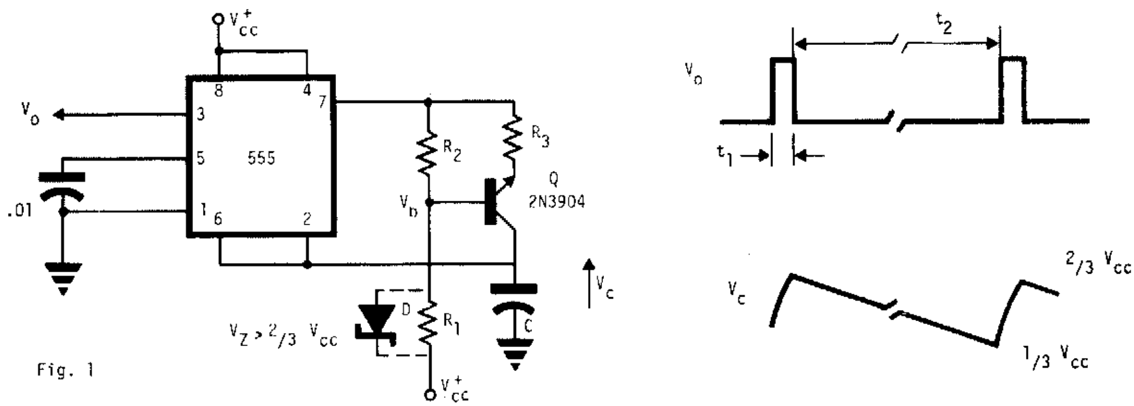
RAMP GENERATOR WITH VARIABLE RESET LEVEL



LINEAR TECHNOLOGY CORP.

Fig. 40-9

555 ASTABLE WITH LOW DUTY CYCLE



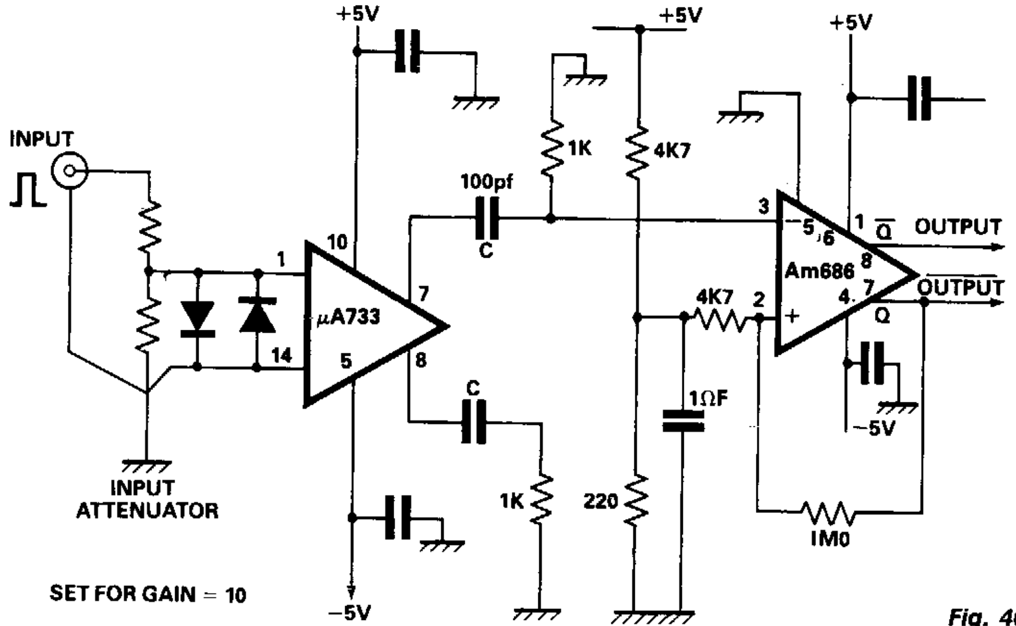
ELECTRONIC ENGINEERING

Fig. 40-10

Circuit Notes

This free-running multivibrator uses an external current sink to discharge the timing capacitor, C. Therefore, interval t_2 may easily be 1000 × the pulse duration, t_1, which defines a positive output. Capacitor voltage, V_C, is a negative going ramp with exponential rise during the pulse output periods.

MONOSTABLE USING VIDEO AMPLIFIER AND COMPARATOR



ELECTRONIC ENGINEERING

Fig. 40-11

Circuit Notes

The output of a video amplifier is differentiated before being fed to a Schottky comparator. The propagation delay is reduced to typically 10ns. The output pulse width is set by the value of C, 100pf giving a pulse of about 90ns duration.

UJT MONOSTABLE CIRCUIT INSENSITIVE TO CHANGE IN BIAS VOLTAGE

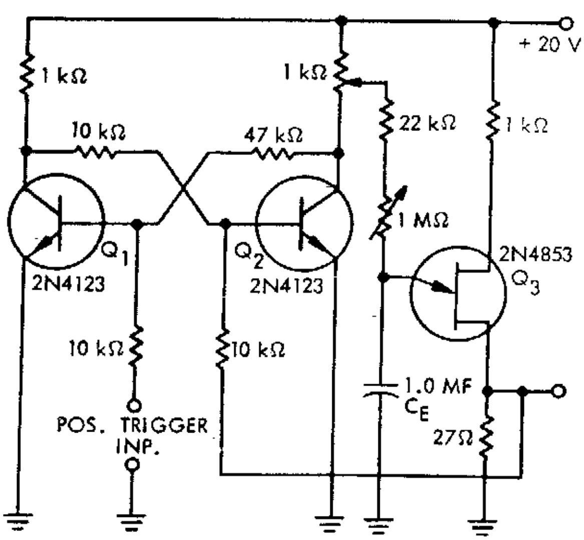


Fig. 40-12

MOTOROLA INC.

ASTABLE MULTIVIBRATOR

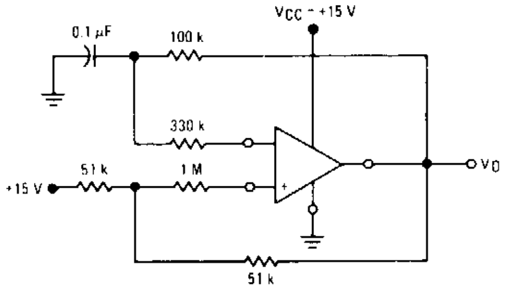
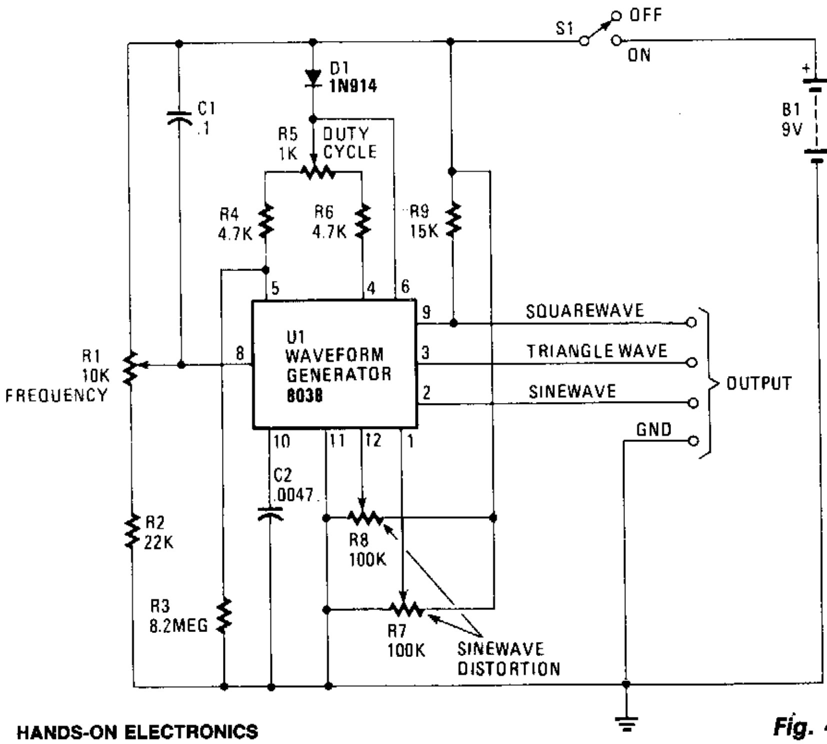


Fig. 40-13

IC = MC3301

MOTOROLA INC.

WAVEFORM GENERATOR



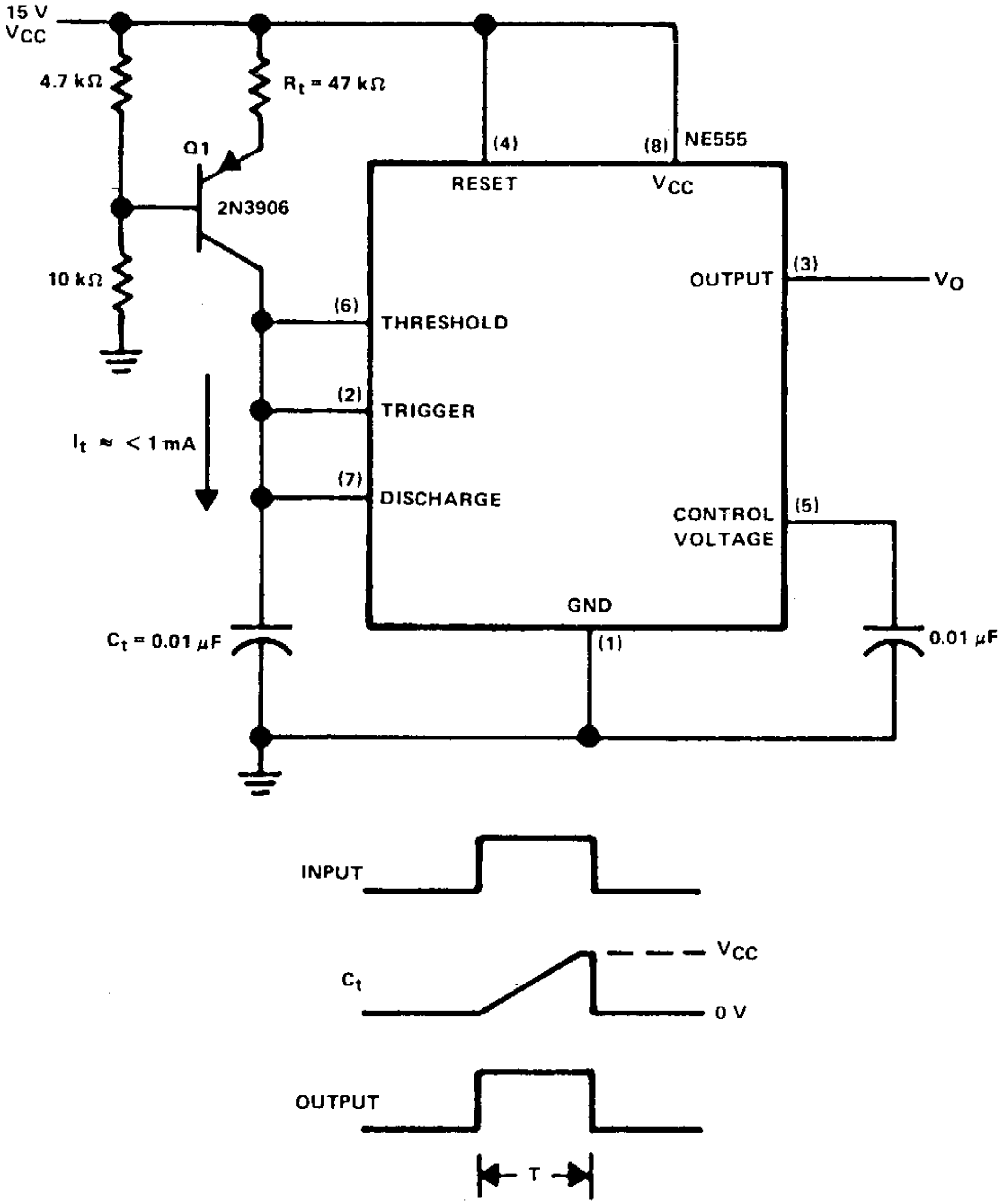
HANDS-ON ELECTRONICS

Fig. 40-14

Circuit Notes

The circuit is designed around the Intersil 8038CC. Frequency range is approximately 20 Hz to 20 kHz—a tuning range of 1000:1 with a single control. The output frequency depends on the value of C2 and on the setting of potentiometer R1. Other values of C2 change the frequency range. Increase the value of C2 to lower the frequency. The lowest possible frequency is around .001 Hz and the highest is around 300 kHz.

LINEAR RAMP GENERATOR



TEXAS INSTRUMENTS

Fig. 40-15

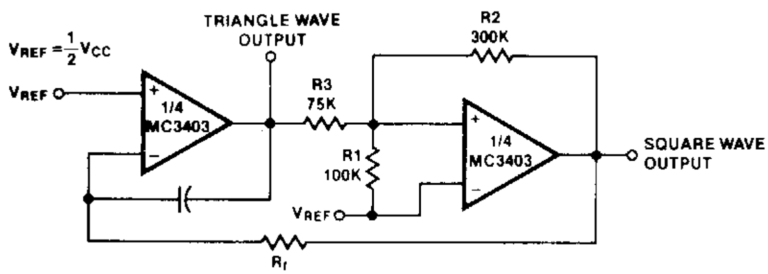
LINEAR RAMP GENERATOR, Continued.

Circuit Notes

The linear charging ramp is most useful where linear control of voltage is required. Some possible applications are a long period voltage controlled timer, a voltage to pulse width converter, or a linear pulse width modulator. Q1 is the current source transistor, supplying constant current to the timing capacitor C_t. When the timer is triggered, the clamp on C_t is removed and C_t charges linearly toward V_{CC} by virtue of the constant current supplied by Q1. The threshold at pin 6 is ⅔ V_{CC}; here, it is termed V_C. When the voltage across C_t reaches V_C volts, the timing cycle ends. The timing expression for output pulse with T is:

In general, I_t should be 1 mA value compatible with the NE555.

FUNCTION GENERATOR

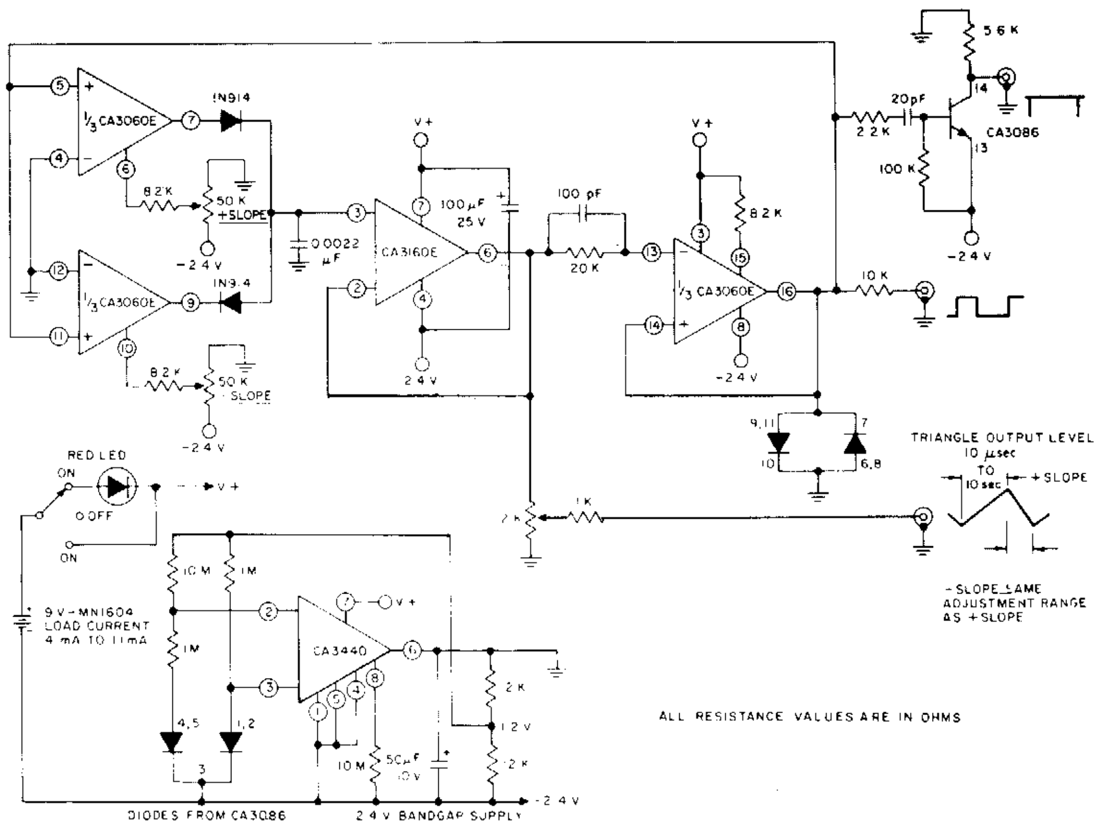


$$f = \frac{R1 + R_C}{4\,C R_f\,R1} \quad \text{IF } R3 = \frac{R2\,R1}{R2 + R1}$$

SIGNETICS

Fig. 40-16

WAVEFORM GENERATOR

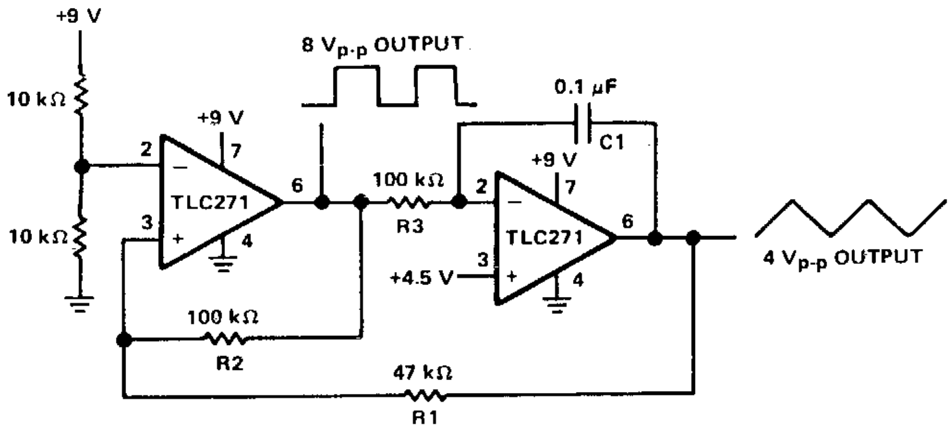


GENERAL ELECTRIC/RCA *Fig. 40-17*

Circuit Notes

The circuit uses a CA3060 triple OTA (two units serve as switched current generators controlled by a third amplifier). A CA3160 BiMOS op amp serves as a voltage follower to buffer the 0.0022 μF integrating capacitor. The circuit has an adjustment range of 1,000,000:1 and a timing range of 20 μs to 20 sec. The "ON-OFF" switch actuates an LED that serves as both a pilot light and a low-battery indicator. The LED extends battery life, since it drops battery voltage to the circuit by approximately 1.2 volts, thus reducing supply current.

SINGLE SUPPLY FUNCTION GENERATOR



TEXAS INSTRUMENTS

Fig. 40-18

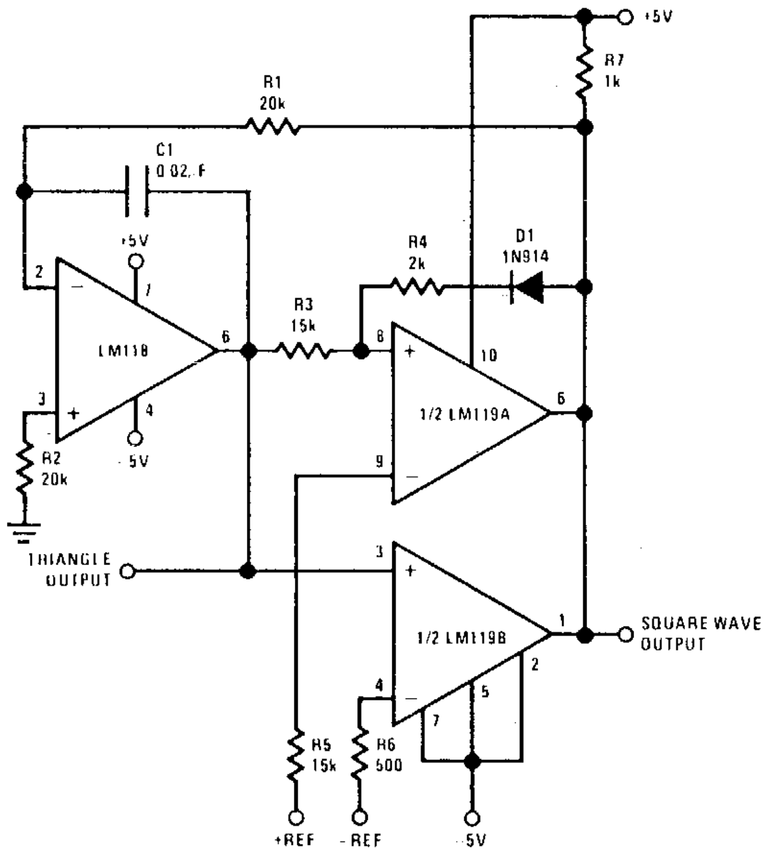
Circuit Notes

The circuit has both square-wave and triangle-wave output. The left section is similar in function to a comparator circuit that uses positive feedback for hysteresis. The inverting input is biased at one-half the V_{CC} voltage by resistors R4 and R5. The output is fed back to the non-inverting input of the first stage to control the frequency. The amplitude of the square wave is the output swing of the first stage, which is 8 V peak-to-peak. The second stage is basically an op amp integrator. The resistor R3 is the input element and capacitor C1 is the feedback element. The ratio R1/R2 sets the amplitude of the triangle wave, as referenced to the square-wave output. For both waveforms, the frequency of oscillation can be determined by the equation:

$$fo = \frac{1}{4R3C1} \quad \frac{R2}{R1}$$

The output frequency is approximately 50 Hz with the given components.

PRECISE WAVE GENERATOR



NATIONAL SEMICONDUCTOR

Fig. 40-19

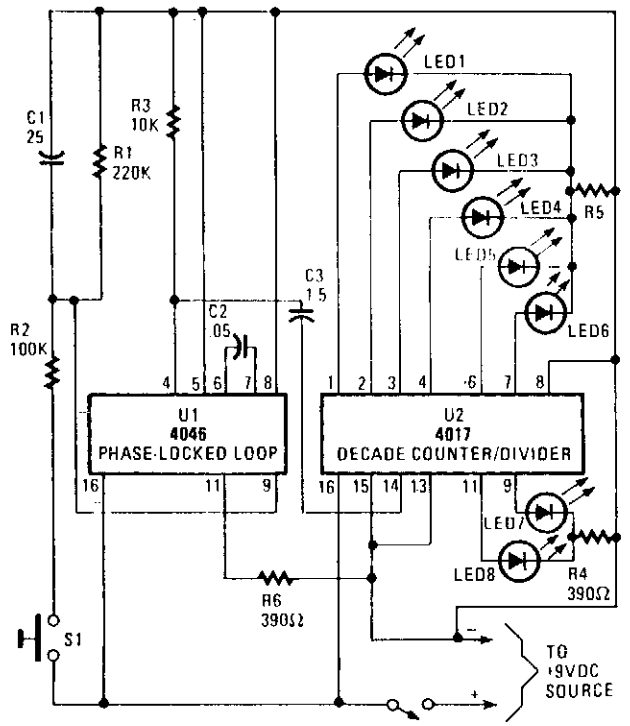
Circuit Notes

The positive and negative peak amplitude is controllable to an accuracy of about ±0.01 V by a dc input. Also, the output frequency and symmetry are easily adjustable. The oscillator consists of an integrator and two comparators—one comparator sets the positive peak and the other the negative peak of the triangle wave. If R1 is replaced by a potentiometer, the frequency can be varied over at least a 10 to 1 range without affecting amplitude. Symmetry is also adjustable by connecting a 50 kΩ resistor from the inverting input of the LM118 to the arm of the 1 kΩ potentiometer. The ends of the potentiometer are connected across the supplies. Current for the resistor either adds or subtracts from the current through R1, changing the ramp time.

41
Games

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ELECTRONIC ROULETTE



TAB BOOKS, INC. *Fig. 41-1*

Circuit Notes

U1 (a 4046 PLL containing a voltage controlled oscillator or VCO, two phase comparators, a source follower, and a Zener diode) is used to produce a low-frequency, pulsed output of about 40 Hz. The VCO's frequency range is determined by R6 and C2, which can be altered by varying the voltage at pin 9. The rising voltage causes the frequency to rise from zero to threshold and remain at that frequency as long as S1 is closed. When S1 is opened, C1 discharges slowly through R1 to ground and the voltage falls toward zero. That produces a decreasing pulse rate. The output of U1 at pin 4 is connected to the clock input of U2 (a 4017 decade decoder/driver) at pin 14 via C3. U2 sequentially advances through each of its ten outputs (0 to 9)—pins 1 to 7, and 9 to 11—with each input pulse. As each output goes high, its associated LED is lighted, and extinguished when it returns to the low state. Only eight outputs are used in the circuit, giving two numbers to the spinner of the house. The circuit can be set up so that the LED's lights sequence or you can use some staggered combination; the LEDs grouped in a straight line or a circle.

LIE DETECTOR

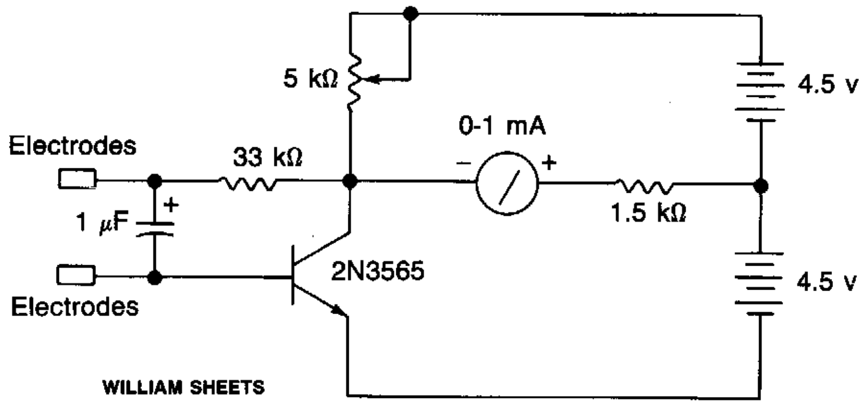


Fig. 41-2

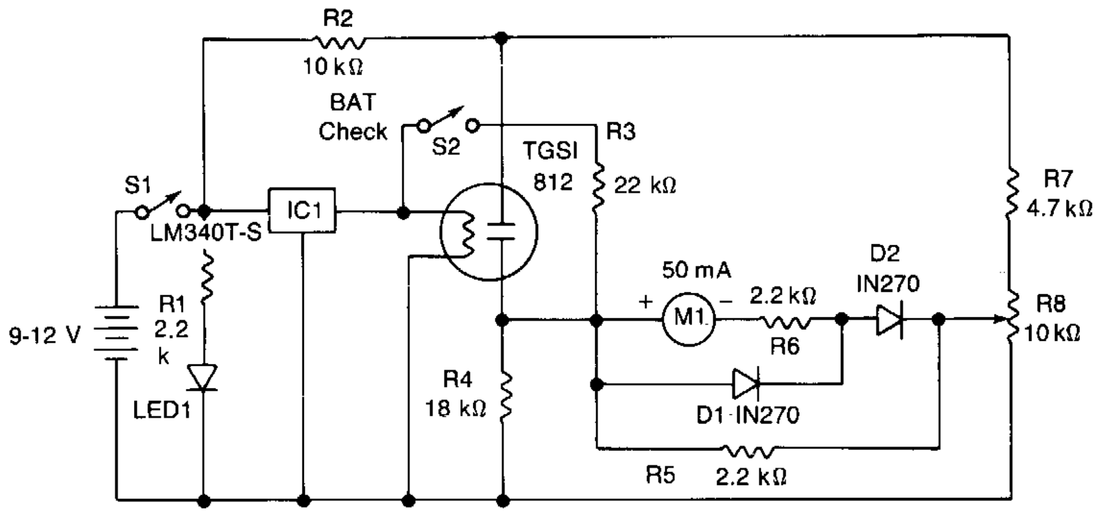
Circuit Notes

The two probes shown are held in the hands and the skin resistance applies bias to the transistor. The 5 k ohm pot is set for zero deflection on the meter. When the "subject" is embarrassed or lies, sweating on the hands takes place, increasing the bias to the transistor and upsetting the bridge balance.

42
Gas and Smoke Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

GAS AND VAPOR DETECTOR



WILLIAM SHEETS

Fig. 42-1

Circuit Notes

The power drain is approximately 150 mA. IC1 provides a regulated 5-volt supply for the filament heater of the sensor. The gas sensitive element is connected as one arm of a resistance bridge consisting of R4, R7, R8 and the meter M1 with its associated resistors R5 and R6. The bridge can be balanced by adjusting R8 so that no current flows through the meter. A change in the sensor's resistance, caused by detection of noxious gases, will unbalance the circuit and deflect the meter. Diodes D1, D2 and resistor R5 protect the meter from overload while R6 determines overall sensitivity. R2 limits the current through the sensor; R1 and LED1 indicate that the circuit is working, so that you do not drain the battery leaving the unit on inadvertently; R3 and S2 give a battery level check.

TOXIC GAS DETECTOR

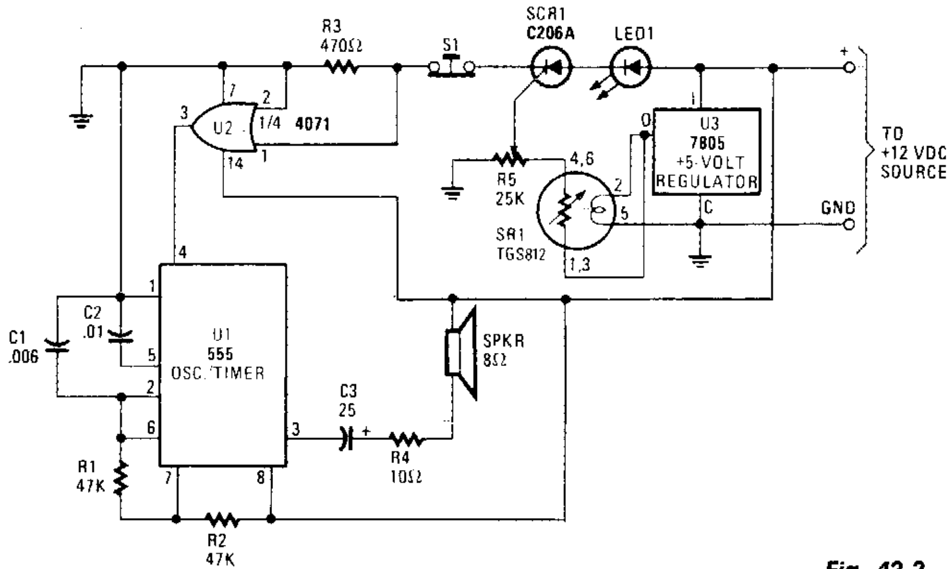


Fig. 42-2

HANDS-ON ELECTRONICS

Circuit Notes

The major device in the circuit is SR1 (a TGS812 toxic-gas sensor manufactured by Figaro Engineering Inc.) The gas-sensitive semiconductor (acting like a variable resistor in the presence of toxic gas) decreases in electrical resistance when gaseous toxins are absorbed from the sensor surface. A 25,000 ohm potentiometer (R5) connected to the sensor serves as a load, voltage-dividing network, and sensitivity control and has its center tap connected to the gate of SCR1. When toxic fumes come in contact with the sensor, decreasing its electrical resistance, current flows through the load (potentiometer R5). The voltage developed across the wiper of R5, which is connected to the gate of SCR1, triggers the SCR into conduction. With SCR1 now conducting, pin 1-volt supply for the semiconductor elements of the TGS812 in spite of the suggested 10 volts, thus reducing the standby current. A 7805 regulator is used to meet the 5-volt requirement for the heater and semiconductor elements.

GAS ANALYZER

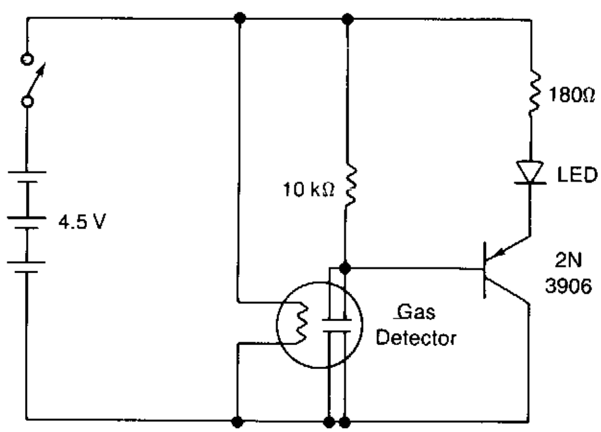


Fig. 42-3

WILLIAM SHEETS

Circuit Notes

The circuit shows a simple yes/no gas detector. Three 1.5-V D cells are used as a power supply, with S1 acting as an on/off switch. The heater is energized directly from the battery, while the electrodes are in series with a 10 k resistor. The voltage across this resistor is monitored by a pnp transistor. When the sensor is in clean air, the resistance between the electrodes is about 40 k, so that only about 0.9 V is dropped across the 10 k resistor. This is insufficient to turn on the transistor, because of the extra 1.6 V required to forward bias the light emitting diode (LED) in series with the emitter. When the sensor comes in contact with contaminated air, the resistance starts to fall, increasing the voltage dropped across the 10 k resistor. When the sensor resistance falls to about 10 k or less, the transistor starts to turn on, current passes through the LED, causing it to emit. The 180 ohm resistor limits the current through the LED to a safe value.

43
Hall Effect Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ANGLE OF ROTATION DETECTOR

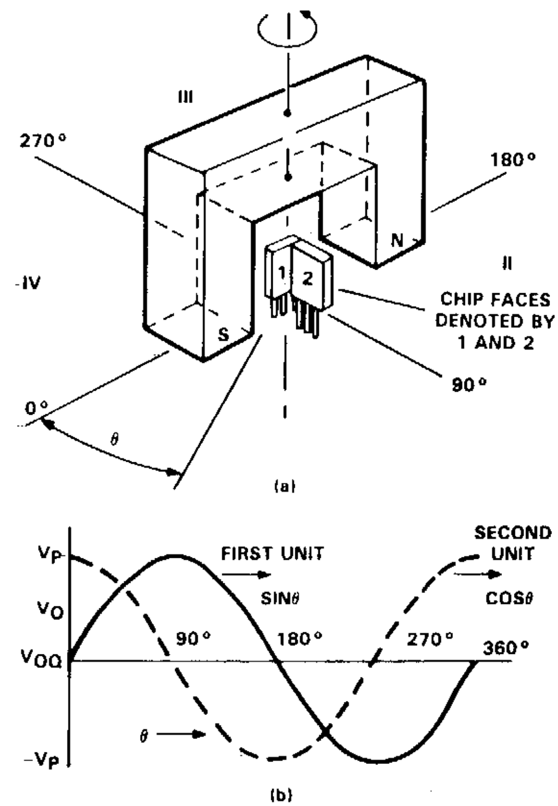


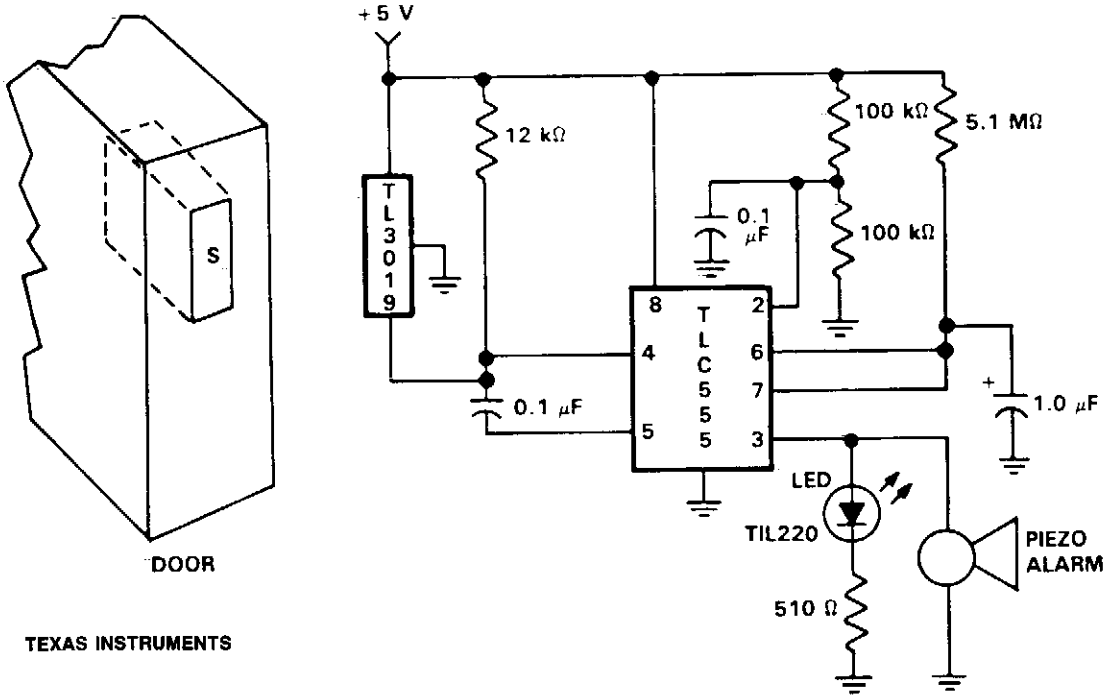
Fig. 43-1

TEXAS INSTRUMENTS

Circuit Notes

The figure shows two TL3103 linear Hall-effect devices used for detecting the angle of rotation. The TL3103s are centered in the gap of a U-shaped permanent magnet. The angle that the south pole makes with the chip face of unit #1 is defined as angle 0. Angle 0 is set to 0° when the chip face of unit #1 is perpendicular to the south pole of the magnet. As the south pole of the magnet sweeps through a 0° to 90° angle, the output of the sensor increases from 0°. Sensor unit #2 decreases from its peak value of +Vp at 0° to a value V_{OQ} at 90°. So, the output of sensor unit #1 is a sine function of 0 and the output of unit #2 is a cosine function of 0 as shown. Thus, the first sensor yields the angle of rotation and the second sensor indicates the quadrant location.

DOOR OPEN ALARM



TEXAS INSTRUMENTS

Fig. 43-2

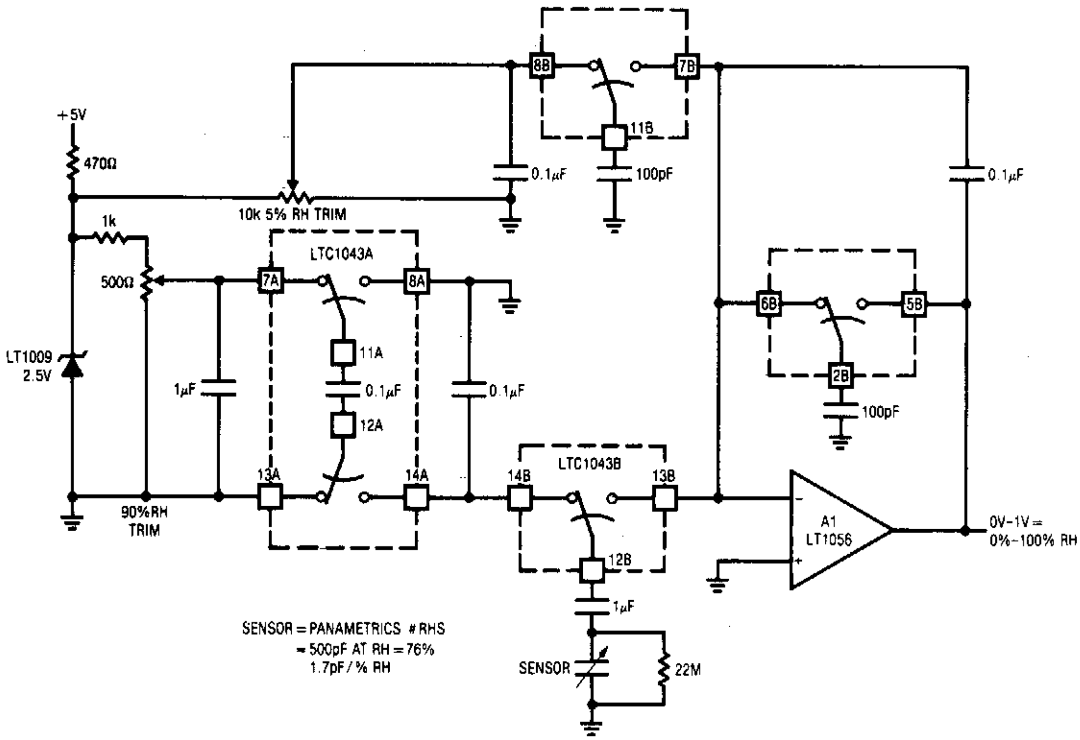
Circuit Notes

Door open alarms are used chiefly in automotive, industrial, and appliance applications. This type of circuit can sense the opening of a refrigerator door. When the door opens, a triac could be activated to control the inside light. The figure shows a door position alarm. When the door is opened, an LED turns on and the piezo alarm sounds for approximately 5 seconds. This circuit uses a TL3019 Hall-effect device for the door sensor. This normally open switch is located in the door frame. The magnet is mounted in the door. When the door is in the closed position, the TL3019 output goes to logic low, and remains low until the door is opened. This design consists of a TLC555 monostable timer circuit. The 1 μF capacitor and 5.1 M ohm resistor on pins 6 and 7 set the monostable RC time constant. These values allow the LED and piezo alarm to remain on about 5 seconds when triggered. One unusual aspect of this circuit is the method of triggering. Usually a 555 timer circuit is triggered by taking the trigger, pin 2, low which produces a high at the output, pin 3. In this configuration with the door in the closed position, the TL3019 output is held low. The trigger, pin 2, is connected to ½ the supply voltage V_{CC}. When the door opens, a positive high pulse is applied to control pin 5 through a 0.1 μF capacitor and also to reset pin 4. This starts the timing cycle. Both the piezo alarm and the LED visual indicator are activated.

44
Humidity Sensors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RELATIVE HUMIDITY SENSOR SIGNAL CONDITIONER

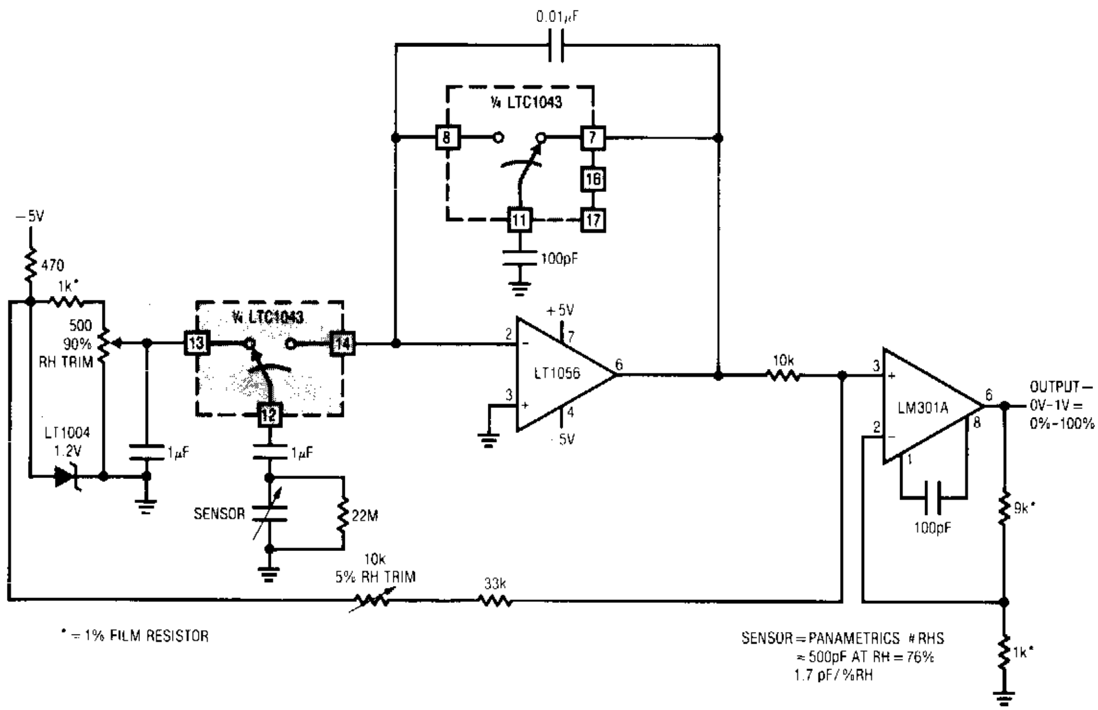


LINEAR TECHNOLOGY CORPORATION

Fig. 44-1

Circuit Notes

This circuit combines two LTC1043s with a based humidity transducer in a simple charge-pump based circuit. The sensor specified has a nominal 400 pF capacitance at RH = 76%, with a slope of 1.7 pF/% RH. The average voltage across this device must be zero. This provision prevents deleterious electrochemical migration in the sensor. The LTC1043A inverts a resistively scaled portion of the LT1009 reference, generating a negative potential at pin 14A. The LTC1043B alternately charges and discharges the humidity sensor via pins 12B, 13B, and 14B. With 14B and 12B connected, the sensor charges via the 1 μF unit to the negative potential at pin 14A. When the 14B-12B pair

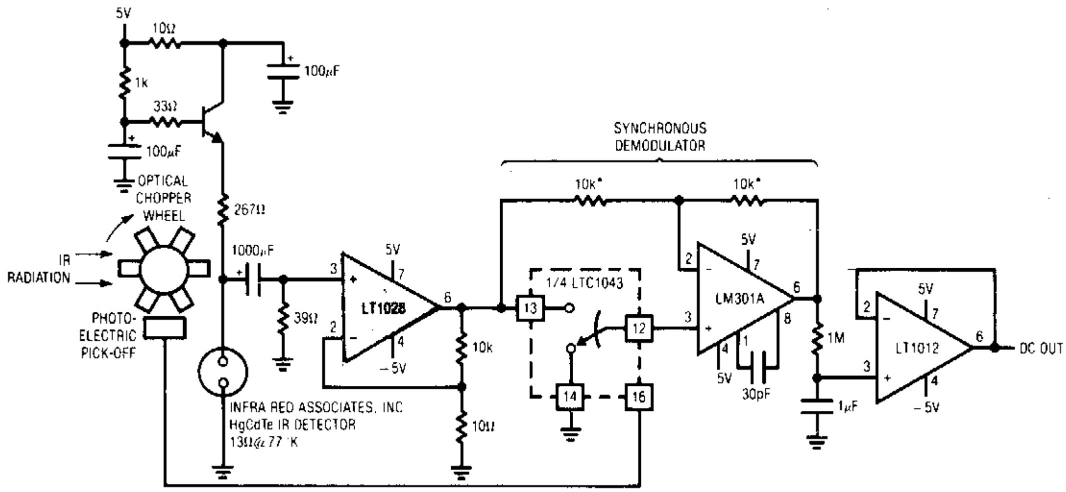


opens, 12B is connected to A1's summing point via 13B. The sensor now discharges into the summing point through the 1 μF capacitor. Since the charge voltage is fixed, the average current into the summing point is determined by the sensor's humidity related value. The 1 μF value ac couples the sensor to the charge-discharge path, maintaining the required zero average voltage across the device. The 22M resistor prevents accumulation of charge, which would stop current flow. The average current into A1's summing point is balanced by packets of charge delivered by the switched-capacitor gives A1 an integrator-like response, and its output is dc.

45
Infrared Circuits

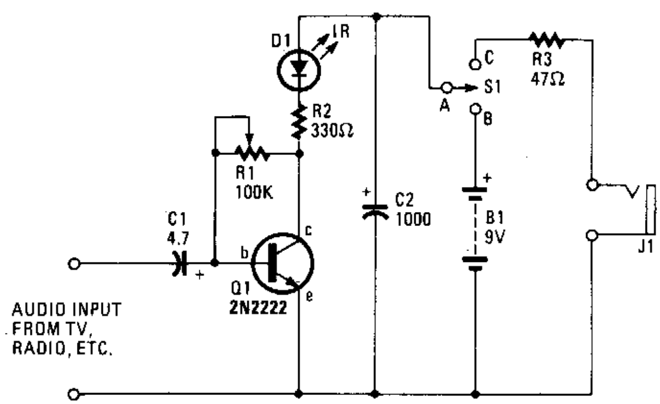
The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW NOISE INFRARED DETECTOR



LINEAR TECHNOLOGY CORP. *Fig. 45-1*

INFRARED TRANSMITTER

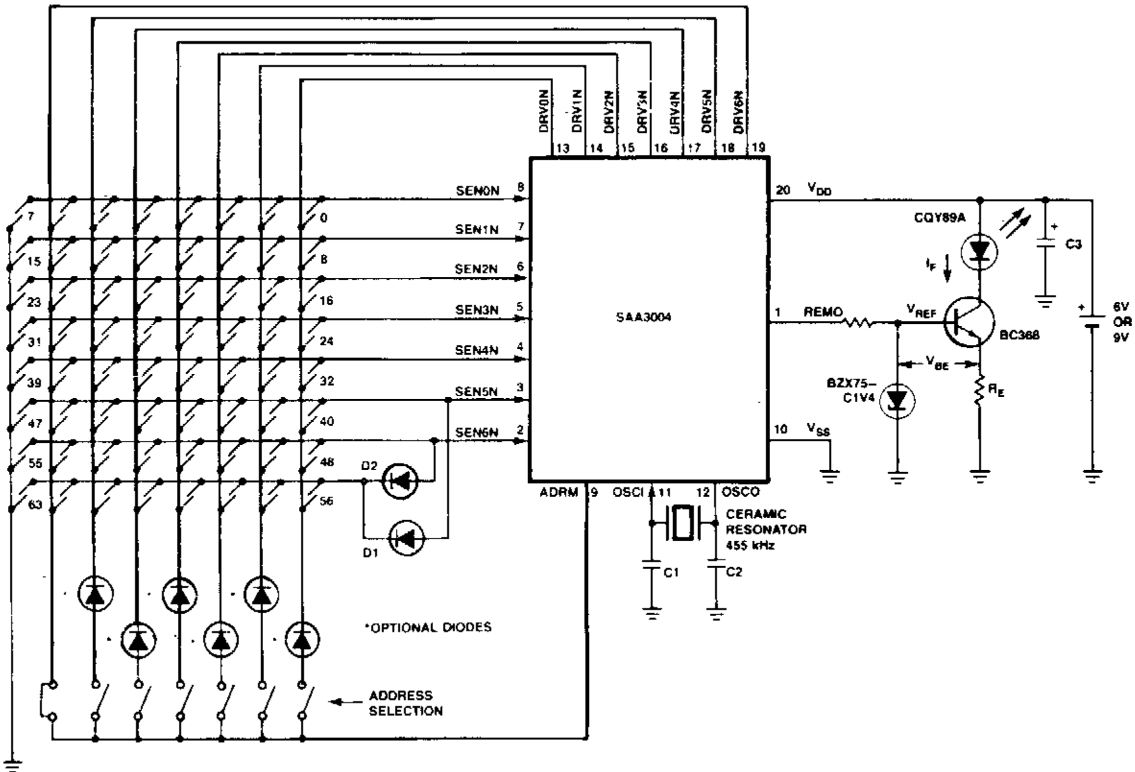


HANDS-ON ELECTRONICS *Fig. 45-2*

Circuit Notes

The ultra-simple one-transistor, IR transmitter shown is designed to transmit the sound from any 8 or 16 ohm audio source, such as a TV, radio, or tape recorder on an infrared beam of light.

INFRARED TRANSMITTER



SIGNETICS

Fig. 45-3

Circuit Notes

The transmitter keyboard is arranged as a scanned matrix. The matrix consists of 7 driver outputs and 7 sense inputs. The driver outputs DRVON to DRV6N are open-drain n-channel transistors and they are conductive in the stand-by mode. The 7 sense inputs (SENON to SEN6N) enable the generation of 56 command codes. With 2 external diodes all 64 commands are addressable. The sense inputs have p-channel pull-up transistors, so that they are HIGH until they are pulled LOW by connecting them to an output via a key depression to initiate a code transmission.

INVISIBLE INFRARED PULSED LASER RIFLE

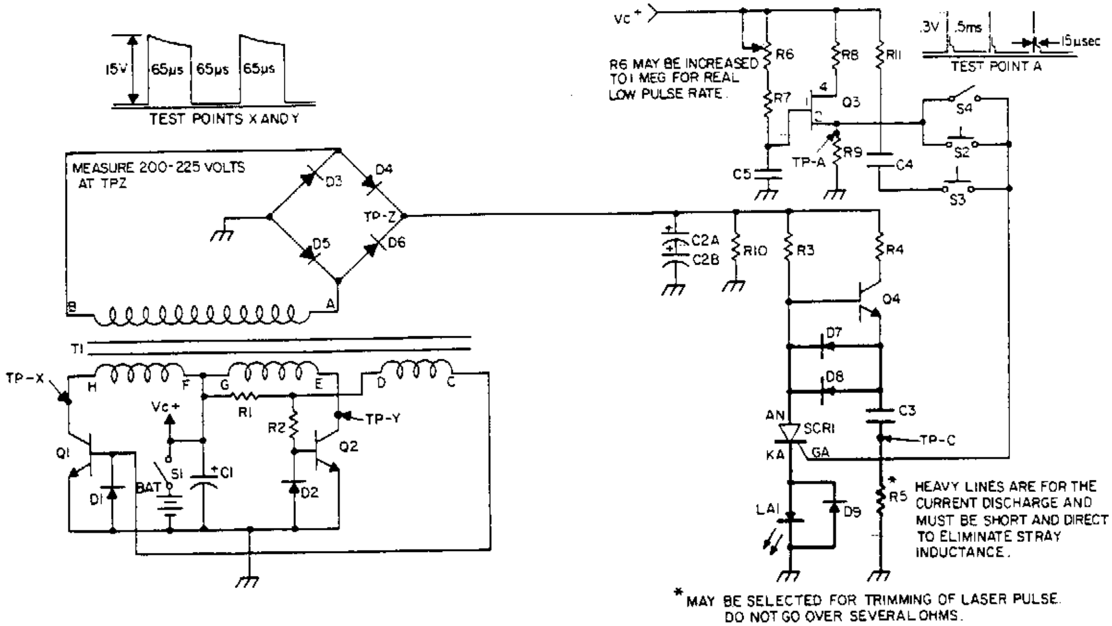


Fig. 45-4

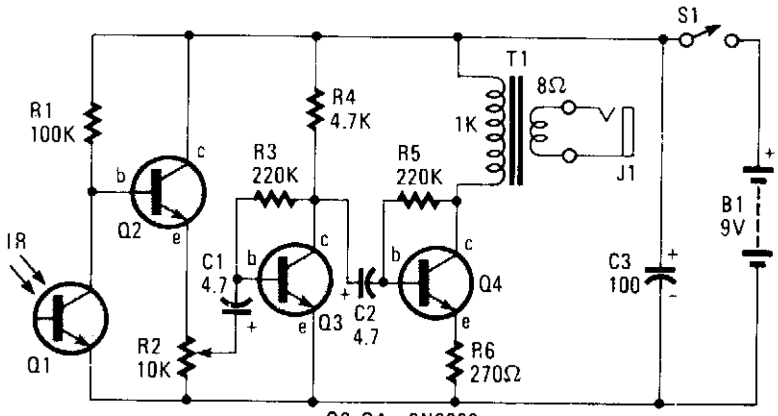
TAB BOOKS INC.

Circuit Notes

The device generates an adjustable frequency of low to medium powered IR pulses of invisible energy and must be treated with care.

The portable battery pack is stepped up to 200 to 300 volts by the inverter circuit consisting of Q1, Q2, and T1. Q1 conducts until saturated, at which time, the base no longer can sustain it in an "on" state and Q1 turns "off," causing the magnetic field in its collector winding to collapse thus producing a voltage or proper phase in the base drive winding that turns on Q2 until saturated, repeating the above sequence of events in an "on/off" action. The diodes connected at the bases provide a return path for the base drive current. The stepped up squarewave voltage on the secondary of T1 is rectified and integrated on C2.

INFRARED RECEIVER



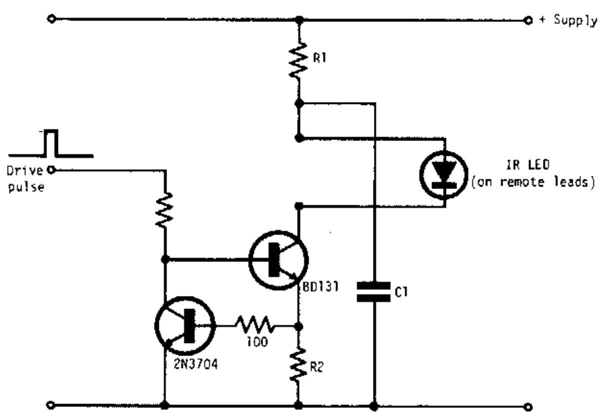
HANDS-ON ELECTRONICS

Fig. 45-5

Circuit Notes

The circuit consists of Q1—a phototransistor that responds to an intensity of amplitude-modulated IR light source—and a three-stage, high-gain audio amplifier. Transformer T1 is used to match the output impedance of the receiver to today's popular low-impedance (low-Z) headphones; but if a set of 1000-2000 ohm, magnetic (not crystal), high-impedance (high-Z) phones are to be used, remove T1 and connect the high-Z phones in place of T1's primary winding—the 1000-ohm winding.

PULSED INFRARED DIODE EMITTER DRIVE



Circuit Notes

Q1 and Q2 form a constant current drive defined by R2. (I approximates to the reciprocal of R2 in the circuit shown for values of I greater than 1 amp). The pulse current is drawn from C1 which is recharged during the time between pulses via R1. The value of C1 is determined from the duration and magnitude of the peak current required, and the time constant R1 C1 is determined from the duration between pulses.

ELECTRONIC ENGINEERING

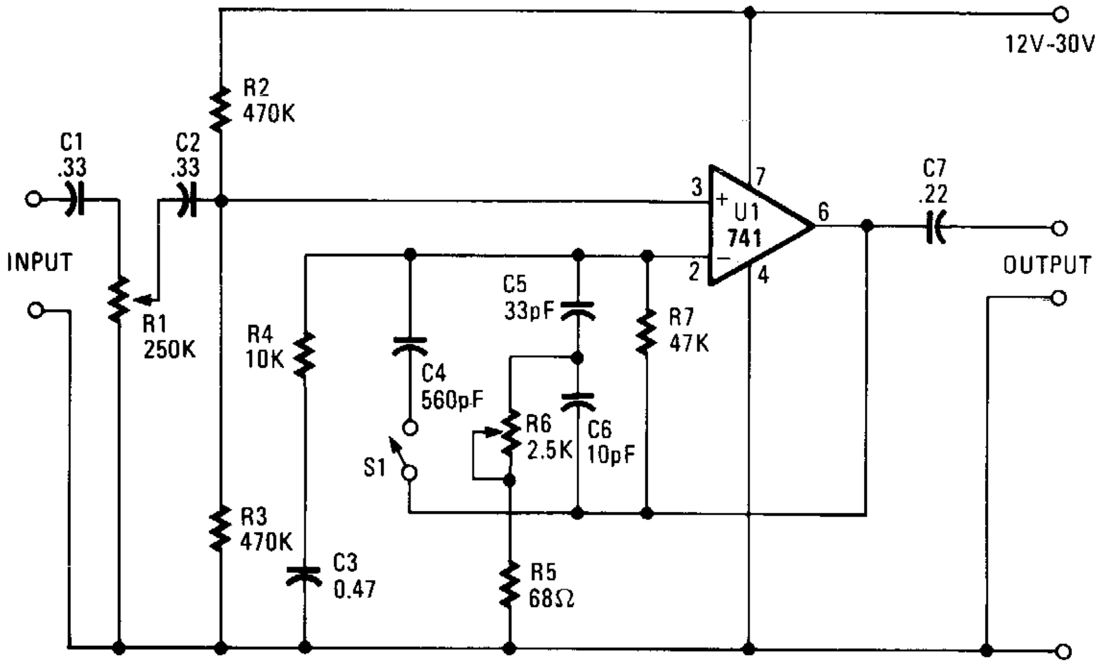
Fig. 45-6

46

Instrumentation Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

INSTRUMENT PREAMP



HANDS-ON ELECTRONICS

Fig. 46-1

Circuit Notes

The input impedance is the value of potentiometer R1. If your instrument has extra-deep bass, change capacitor C1 to 0.5 μF. What appears to be an extra part in the feedback loop is a brightening tone control. The basic feedback from the op amp's output (pin 6) to the inverting input (pin 2) consists of resistor R7, and the series connection of resistor R4 and capacitor C3, which produce a voltage gain of almost 5 (almost 14 dB). That should be more extra oomph than usually needed. If the circuit is somewhat short on bass response, increase the value of capacitor C3 to 1 to 10 μF. Start with 1 μF and increase the value until you get the bass effect you want.

INSTRUMENTATION AMPLIFIER WITH ±100 VOLT COMMON MODE RANGE

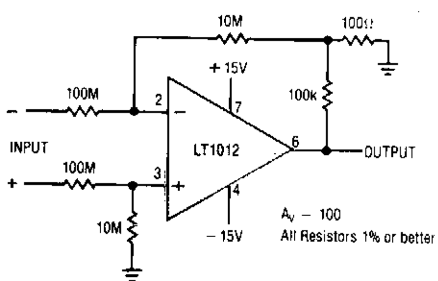
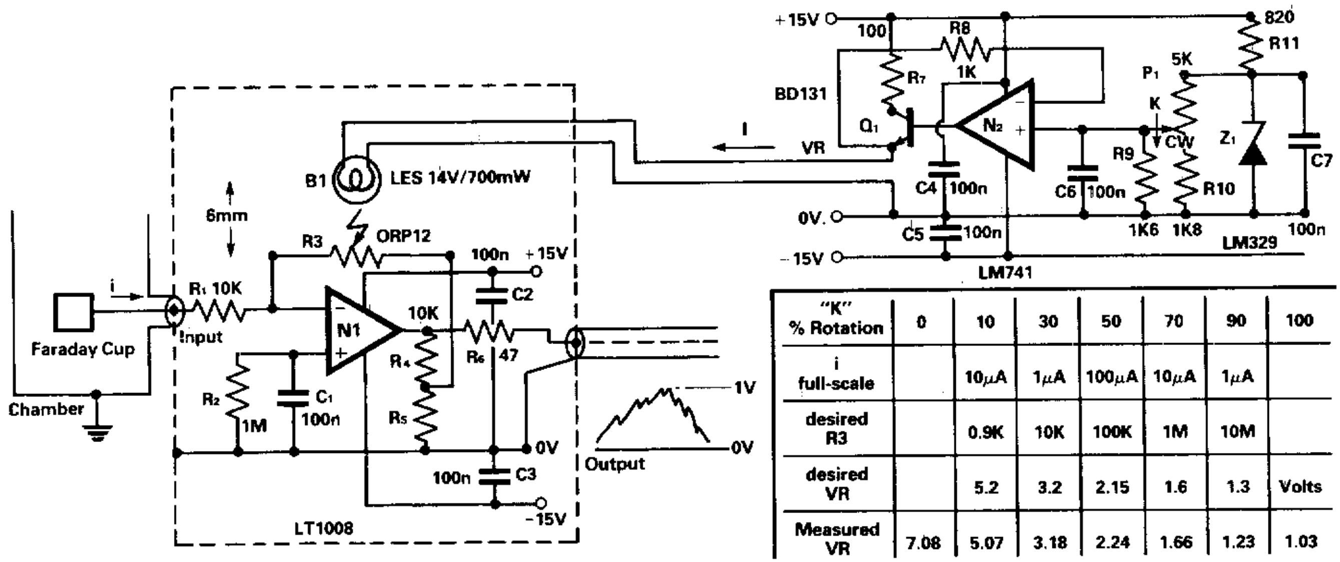


Fig. 46-2

LINEAR TECHNOLOGY CORP.

CURRENT-COLLECTOR HEAD-AMPLIFIER



"K" % Rotation	0	10	30	50	70	90	100
i full-scale		10μA	1μA	100μA	10μA	1μA	
desired R3		0.9K	10K	100K	1M	10M	
desired VR		5.2	3.2	2.15	1.6	1.3	Volts
Measured VR	7.08	5.07	3.18	2.24	1.66	1.23	1.03

ELECTRONIC ENGINEERING

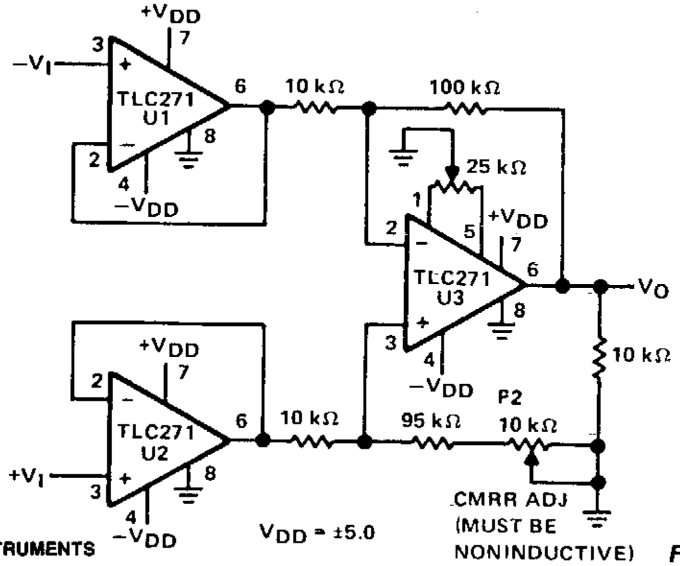
Fig. 46-3

Circuit Notes

To amplify small current signals such as from an electron-collector inside a vacuum chamber, it is convenient for reasons of noise and bandwidth to have a "head-amplifier" attached to the chamber. The op amp N1 is a precision bipolar device with extremely low bias current and offset voltage (1) as well as low noise, which allows the 100:1 feedback attenuator R4:R5. The resistance of R3 can be varied from above 10M to below 1R, and so the nominal 0 to 1 V-peak output signal corresponds to input current ranges of 1 nA to 10 μA; this current i enters via the protective resistor R1. Light from the bulb B1 shines on R3, and the filament current I is controlled by the op amp N2.

The reference voltage VR is "shaped" by the resistors R9R10 so as to tailor the bulb and LDR characteristics to the desired current ranges. Thus, rotation of the calibrated knob K gives the appropriate resistance to R3 for the peak-current scale shown.

INSTRUMENTATION METER DRIVER



TEXAS INSTRUMENTS

Fig. 46-4

Circuit Notes

Three op amps U1, U2, and U3 are connected in the basic instrumentation amplifier configuration. Operating from ±5 V, pin 8 of each op amp is connected directly to ground and provides the ac performance desired in this application (high bias mode). P1 is for offset error correction and P2 allows adjustment of the input common mode rejection ratio. The high input impedance allows megohms without loading. The resulting circuit frequency response is 200 kHz at −3 dB and has a slew rate of 4.5 V/μs.

SATURATED STANDARD CELL AMPLIFIER

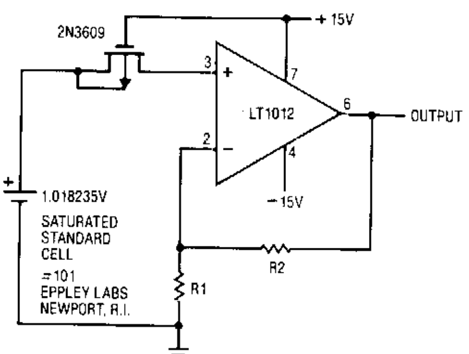


Fig. 46-5

The typical 30pA bias current of the LT1012 will degrade the standard cell by only 1 ppm/year. Noise is a fraction of a ppm. Unprotected gate MOSFET isolates standard cell on power down.

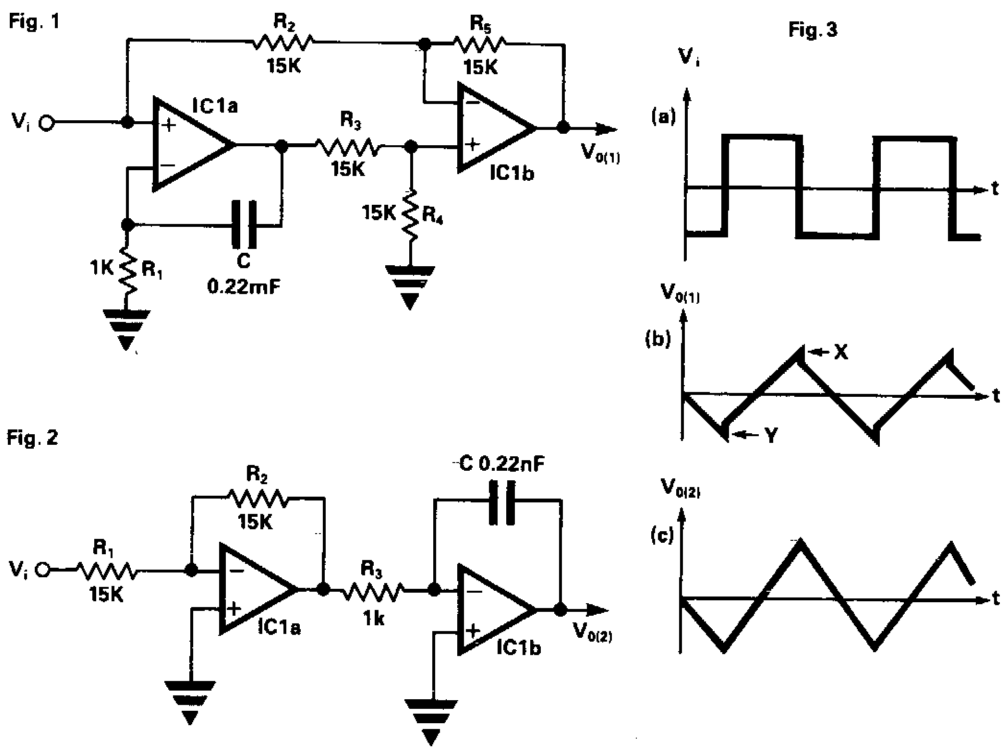
LINEAR TECHNOLOGY CORP.

47

Integrator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

IMPROVED NON-INVERTING INTEGRATOR



ELECTRONIC ENGINEERING

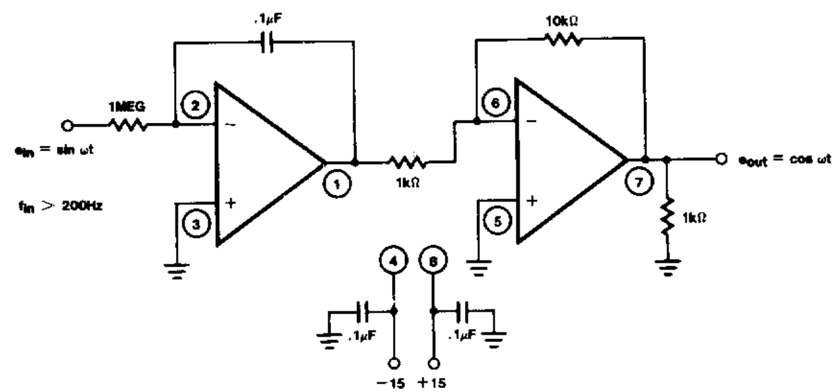
Fig. 47-1

Circuit Notes

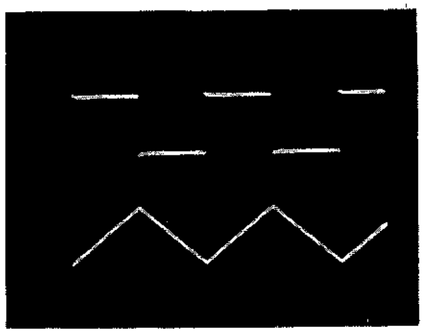
In the circuit in Fig. 1, IC1a produces the integral term required but also has the side effect of producing a proportional term not required, so this term is subtracted by IC1b leaving a pure integral. If the ratio R2/R5 does not exactly match the ratio of R3/R4, the subtraction will not be complete and a small amount of the proportional term will reach the output. The result of this with a squarewave input is shown in Fig. 3a as small steps in the output waveform at points X and Y.

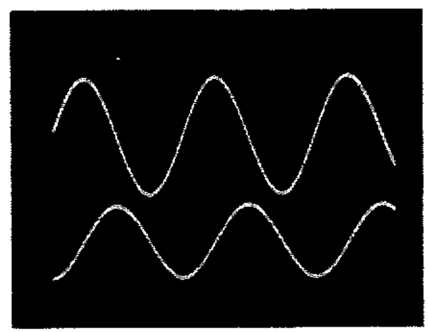
This effect can be completely removed by using the simplified circuit shown in Fig. 2. Here the signal is pre-inverted by IC1a, then fed to a standard inverting integrator IC1b. The result is a non-inverting integrator with the advantage that the unwanted proportional term is never produced, so it does not need to be subtracted.

ACTIVE INTEGRATOR WITH INVERTING BUFFER



INTEGRATOR WAVEFORMS





IC = NE/SE5512

SIGNETICS

Fig. 47-2

LONG TIME INTEGRATOR

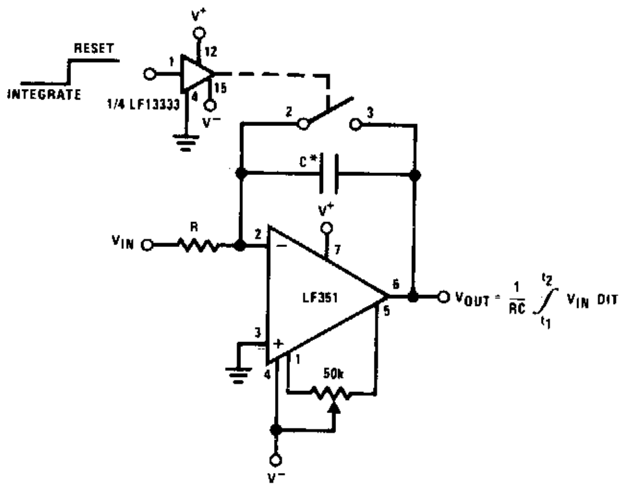


Fig. 47-3

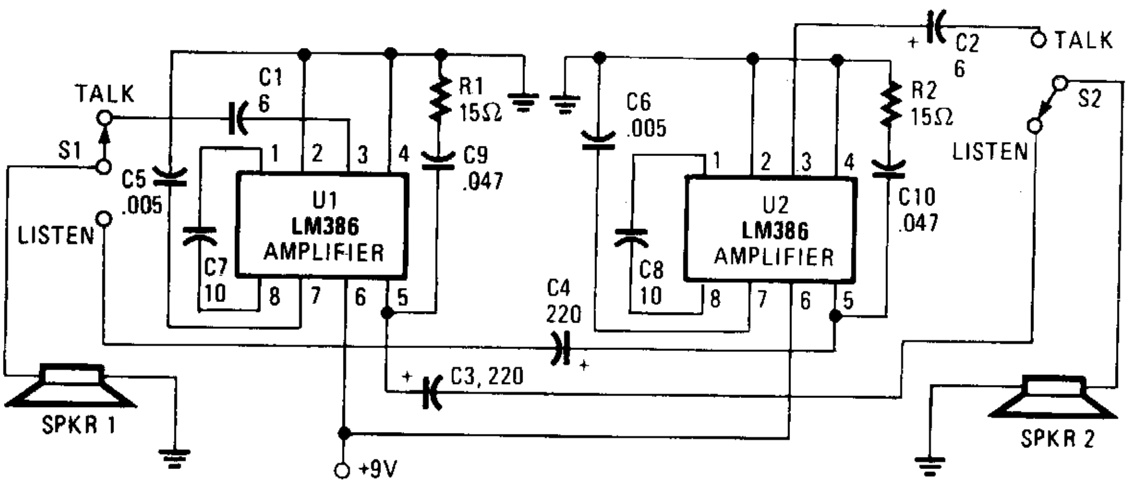
* Low leakage capacitor
• 50k pot used for less sensitive V_{OS} adjust

NATIONAL SEMICONDUCTOR CORP.

48
Intercom Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

INTERCOM



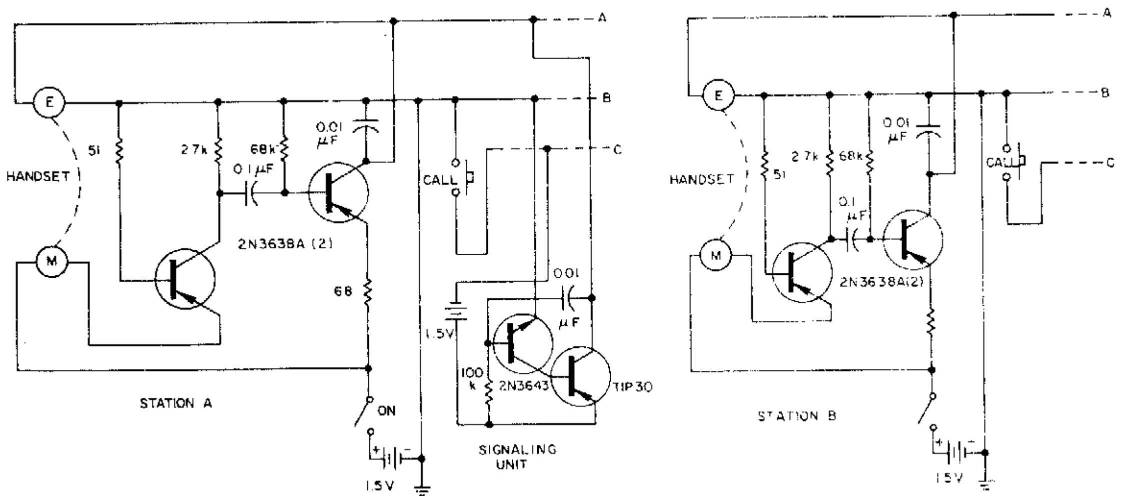
HANDS-ON ELECTRONICS

Fig. 48-1

Circuit Notes

The circuit consists of separate amplifiers—one for each station—rather than a single amplifier and a time sharing arrangement. U1 and U2 are low-voltage audio amplifiers, each of which operates as separate entities with switches at either station controlling which will transmit or receive. With capacitors C7 and C8 included in the circuit, the amplifiers have a gain of 200. Omitting those two components drops the gain to about 20. Other gain levels are available with the addition of a series-connected R/C combination connected between pin 1 and pin 8—for example, a 1000 ohm resistor and 10 μF capacitor for a gain of about 150.

PARTY-LINE INTERCOM



ELECTRONIC DESIGN

Fig. 48-2

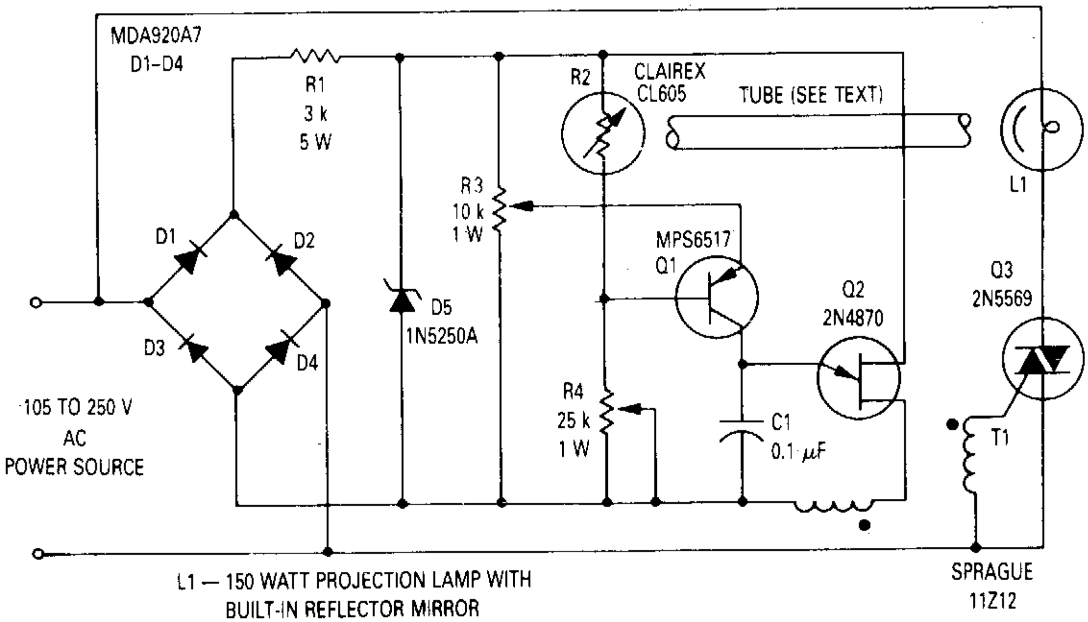
Circuit Notes

A large number of intercom stations can be tied together. All units are connected in parallel, and the entire system is buzzed by only one signaling circuit. Each unit is powered individually from 1.5-V cells for redundancy. For greater signal volume, 3-V sources can be used for the supplies without changing any other parts of the system. The carbon microphone of a standard telephone handset at each station feeds into a common-base amplifier, and a tandem high-gain common-emitter stage drives the intercom line. All phone earpieces are in parallel across the line. The signaling circuit, also connected across the line, is a simple oscillator that drives all the earpieces.

49
Lamp-Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VOLTAGE REGULATOR FOR A PROJECTION LAMP



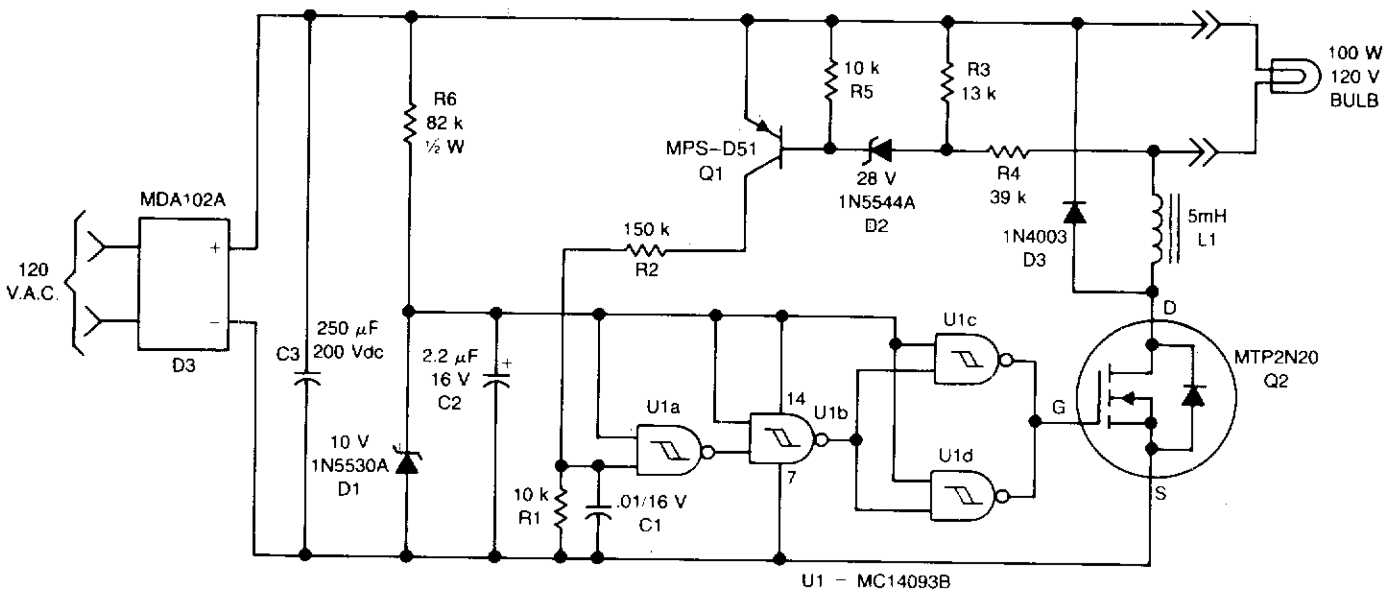
MOTOROLA

Fig. 49-1

Circuit Notes

The circuit will regulate the rms output voltage across the load (a projection lamp) to 100 volts ±2% for an input voltage between 105 and 250 volts ac. This is accomplished by indirectly sensing the light output of lamp L1 and applying this feedback signal to the firing circuit (Q1 and Q2) which controls the conduction angle of TRIAC Q3. The lamp voltage is provided by TRIAC Q3, whose conduction angle is set by the firing circuit for unijunction transistor Q2. The circuit is synchronized with the line through the full-wave bridge rectifier. The voltage to the firing circuit is limited by zener diode D5. Phase control of the supply voltage is set by the charging rate of capacitor C1. Q2 will fire when the voltage on C1 reaches approximately 0.65 times the zener voltage. The charging rate of C1 is set by the conduction of Q1, which is controlled by the resistance of photocell R2. Potentiometers R3 and R4 are used to set the lamp voltage to 100 volts when the line voltage is 105 volts and 250 volts, respectively.

MACHINE VISION ILLUMINATION STABILIZER



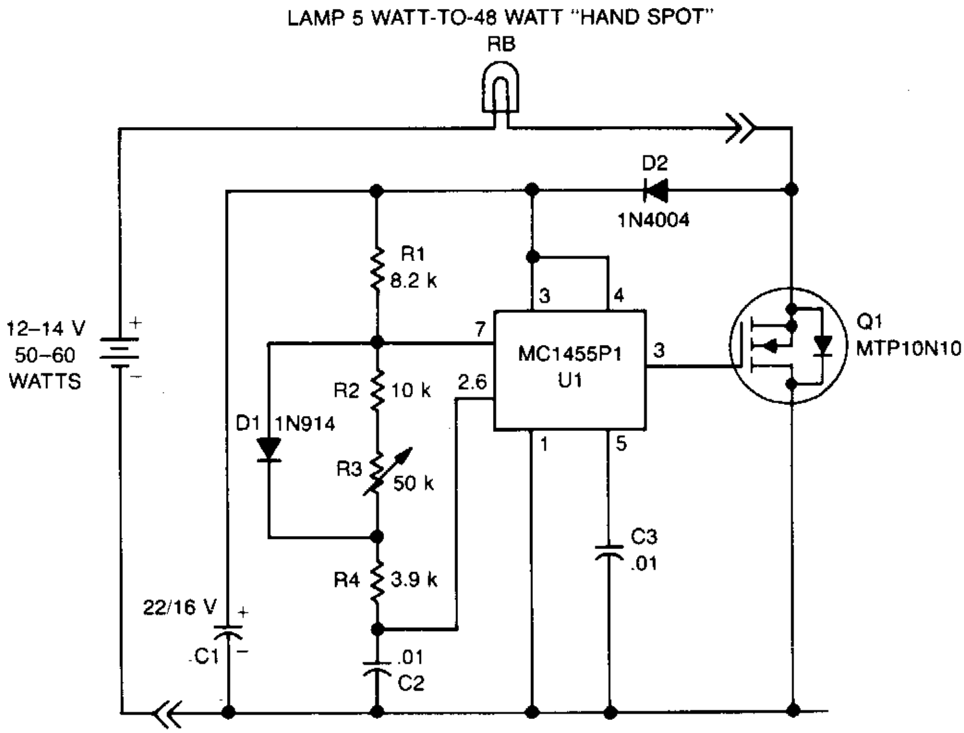
MOTOROLA

Fig. 49-2

Circuit Notes

The combination of Q1, Q2 and U1 form a hysteresis oscillator to stabilize lamp illumination. In operation, full wave bridge D3 operates directly from the ac line to supply unregulated dc to the lamp and also to the 10 V zener that provides power to the quad CMOS Schmitt trigger, U1. When the lamp supply exceeds 115 V, Q1 is turned ON, charging C1 through R2 to raise the input to U1a past the positive-going logic threshold. This drops the output voltage at U1c and U1d, which drives the gate of Q2, turning it OFF. Current then decays through the lamp, L1 and D3 until the lamp voltage falls below 115 V, at which time Q1 turns OFF and the cycle repeats.

DC LAMP DIMMER



MOTOROLA

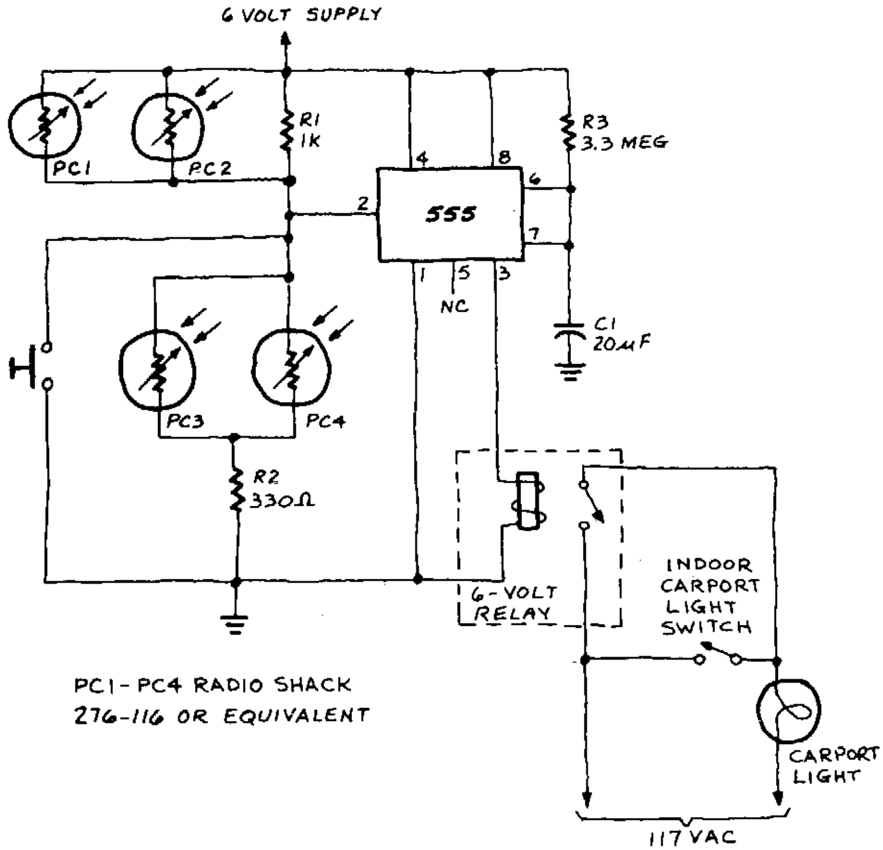
Fig. 49-3

Circuit Notes

A low power, low cost dc lamp dimmer for a two-wire portable "flashlight" can be realized with little or no heatsinking. In addition, a single potentiometer, R3 adjusts lamp brightness.

Battery power is stored in C1 for U1, which is a free-running multivibrator whose frequency is determined by R1, R2, R3, R4, and C2. U1 drives the gate of Q1, turning it and the lamp ON and OFF at a rate proportional to the multivibrator duty cycle.

AUTOMATIC LIGHT CONTROLLER FOR CARPORT

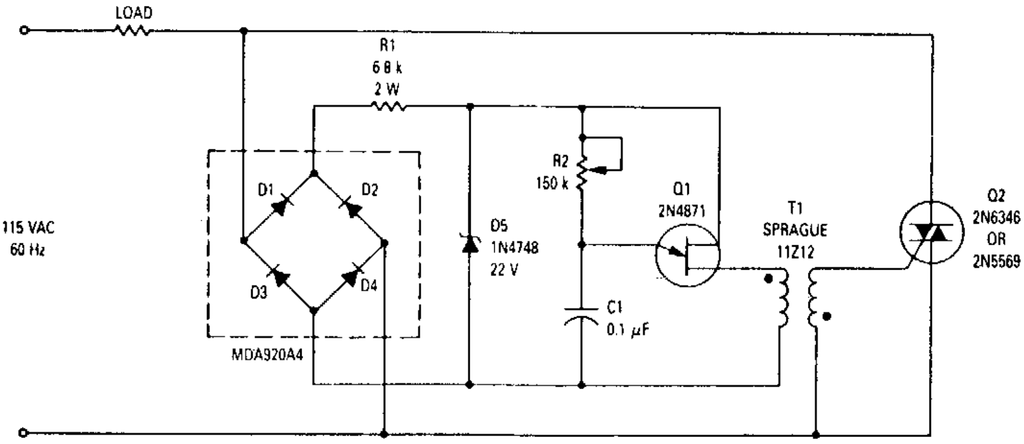


RADIO-ELECTRONICS ***Fig. 49-4***

Circuit Notes

A 555 timer IC, operating in the one-shot mode, is triggered by light striking photoresistors. These normally have a resistance of several megohms but, in the presence of light, that resistance drops to several hundred ohms, permitting current from the six-volt source to flow in the circuit. The R-C combination shown gives an on-time of about two minutes. Photoresistors PC3 and PC4 are mounted at heatlight-height. When headlights illuminate the photoresistor, the timer starts. That actuates a relay, RY1, and the lights are turned on. The lights are automatically turned off when the timer's two minutes are up.

800 W LIGHT-DIMMER



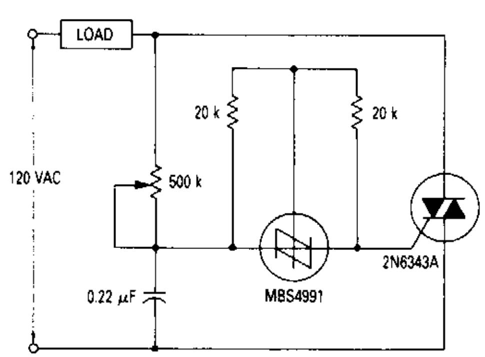
MOTOROLA

Fig. 49-5

Circuit Notes

This wide-range light dimmer circuit uses a unijunction transistor and a pulse transformer to provide phase control for the TRIAC. The circuit operates from a 115 volt, 60 Hz source and can control up to 800 watts of power to incandescent lights. The power to the lights is controlled by varying the conduction angle of the TRIAC from 0° to about 170°. The power available at 170° conduction is better than 97% of that at the full 180°.

LAMP DIMMER



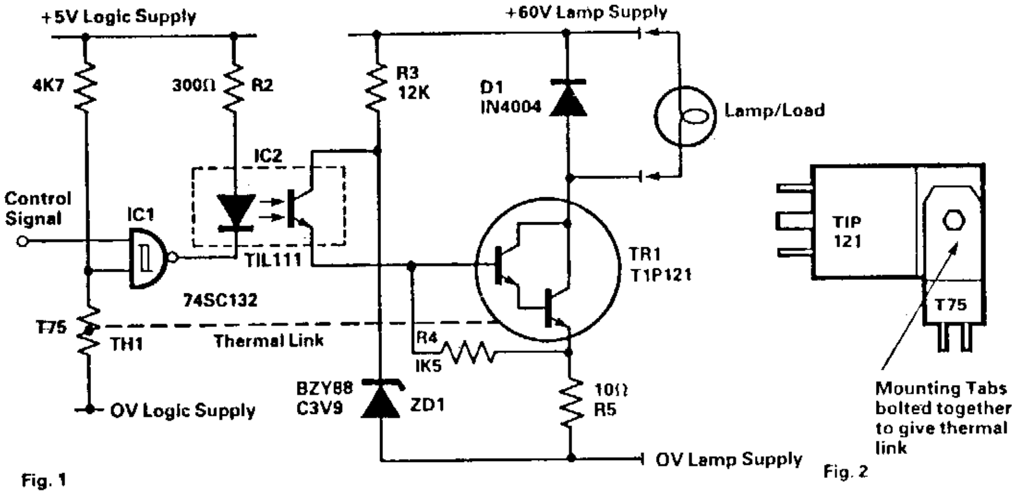
MOTOROLA

Fig. 49-6

Circuit Notes

A full range power controller suitable for lamp dimming and similar applications operate from a 120 volt, 60 Hz ac source, and can control up to 1000 watts of power to incandescent bulbs. The power to the bulbs is varied by controlling the conduction angle of TRIAC Q1. At the end of each positive half-cycle when the applied voltage drops below that of the capacitor, gate current flows out of the SBS and it switches on, discharging the capacitor to near zero volts. The RC network shown across the TRIAC represents a typical snubber circuit that is normally adequate to prevent line transients from accidentally firing the TRIAC.

RUGGED LAMP DRIVER IS SHORT-CIRCUIT PROOF



ELECTRONIC ENGINEERING

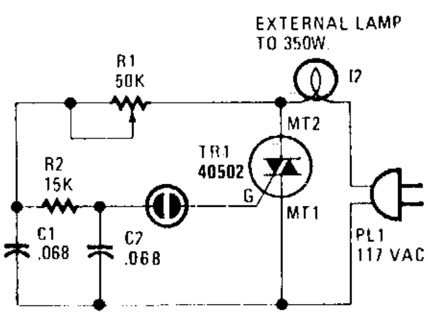
Fig. 49-7

Circuit Notes

This circuit is capable of driving filament lamps of nominal rating 200 mA at 60 V dc from a CMOS logic signal.

The lamp or load is connected in series with the Darlington transistor TR1 and emitter resistor R5. The Zener diode ZD1 establishes a soft reference voltage on the collector of the optical coupler IC2. When the logic control signal from the processor switches the optocoupler on via IC1, base drive is applied to TR1 and the lamp is switched on.

TRIAC LAMP DIMMER



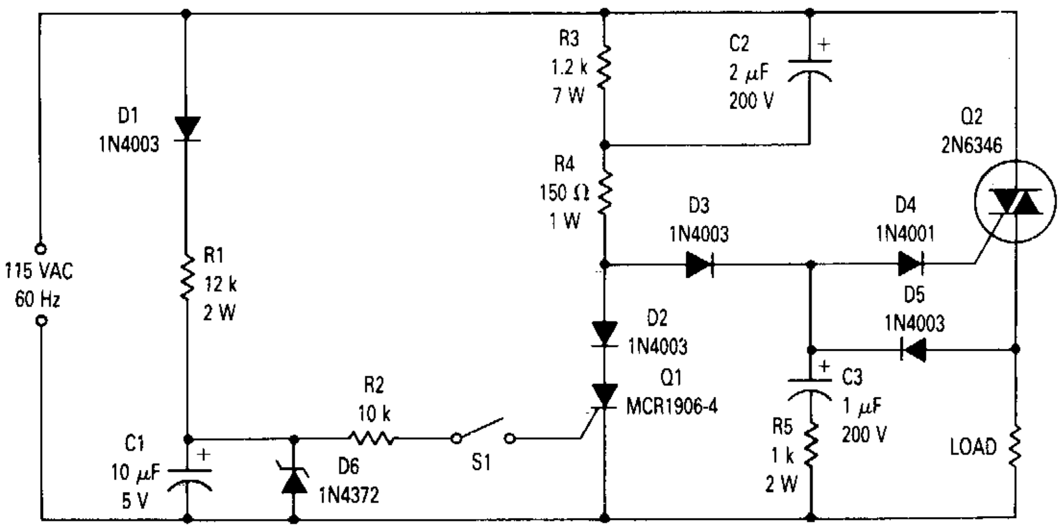
TAB BOOKS, INC.

Fig. 49-8

Circuit Notes

Using a heatsink, the TRIAC (TR1) can handle up to 350 watts. The neon lamp, I1, won't trip the gate until after it conducts and using R1, set the lighting wherever you want it.

TRIAC ZERO-POINT SWITCH



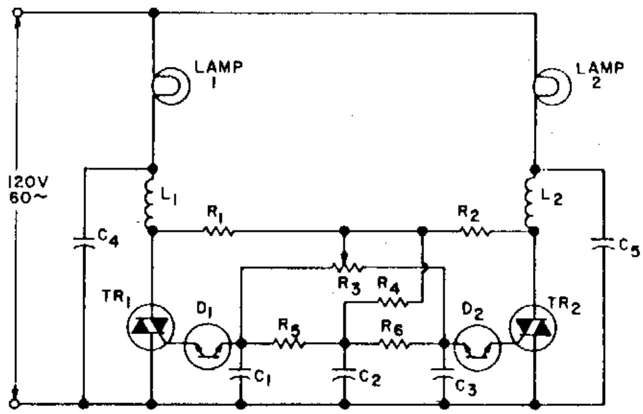
MOTOROLA

Fig. 49-9

Circuit Notes

On the initial part of the positive half cycle, the voltage is changing rapidly from zero causing a large current to flow into capacitor C2. The current through C2 flows through R4, D3, and D4 into the gate of the TRIAC Q2 causing it to turn on very close to zero voltage. Once Q2 turns on, capacitor C3 charges to the peak of the line voltage through D5. When the line voltage passes through the peak, D5 becomes reverse-biased and C3 begins to discharge through D4 and the gate of Q2. At this time the voltage on C3 lags the line voltage. When the line voltage goes through zero there is still some charge on C3 so that when the line voltage starts negative C3 is still discharging into the gate of Q2. Thus Q2 is also turned on near zero on the negative half cycle. This operation continues for each cycle until switch S1 is closed, at which time SCR Q1 is turned on. Q1 shunts the gate current away from Q2 during each positive half cycle keeping Q2 from turning on. Q2 cannot turn on during the negative cycle because C3 cannot charge unless Q2 is on during the positive half cycle.

TANDEM DIMMER (CROSS-FADER)



$R_1 = R_2 = 6800\Omega$, 1 WATT
$R_3 = 150K\ \Omega$ LINEAR POT. 1W
$R_5 = R_6 = 22K\Omega$, 1/2W.
$R_4 = 15K\Omega$, 1/2W.
$TR_1 = TR_2$ = TRIAC
$D_1 = D_2$ = GE ST-2 DIAC

$L_1 = L_2 = 60\mu hy$ (FERRITE CORE)
$C_1 = C_2 = C_3 = 0.1\mu f$ 50V
$C_4 = C_5 = 0.1\mu f$ 200 VOLTS

NOTE: TOTAL LIGHT LEVEL (SUM OF LAMPS 1+2) CONSTANT WITHIN 15%.

GENERAL ELECTRIC

Fig. 49-10

Circuit Notes

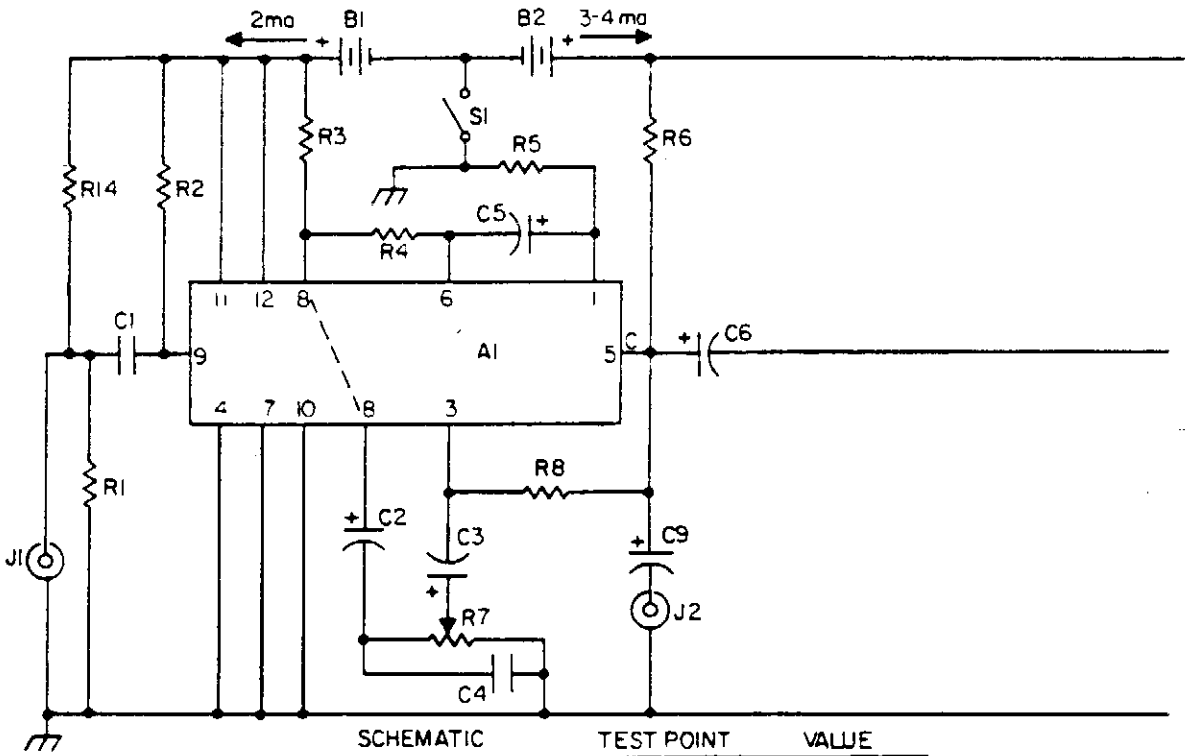
This cross fader circuit can be used for fading between two slide projectors. As R_3 is moved to either side of center, one triac is fired earlier in each half cycle, and the other later. The total light output of both lamps stays about the same for any control position.

50

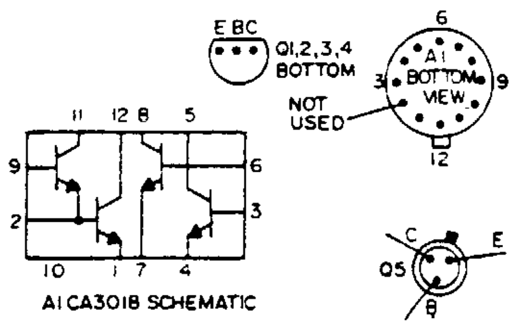
Laser Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LASER LIGHT DETECTOR



SCHEMATIC



A1 CA3018 SCHEMATIC

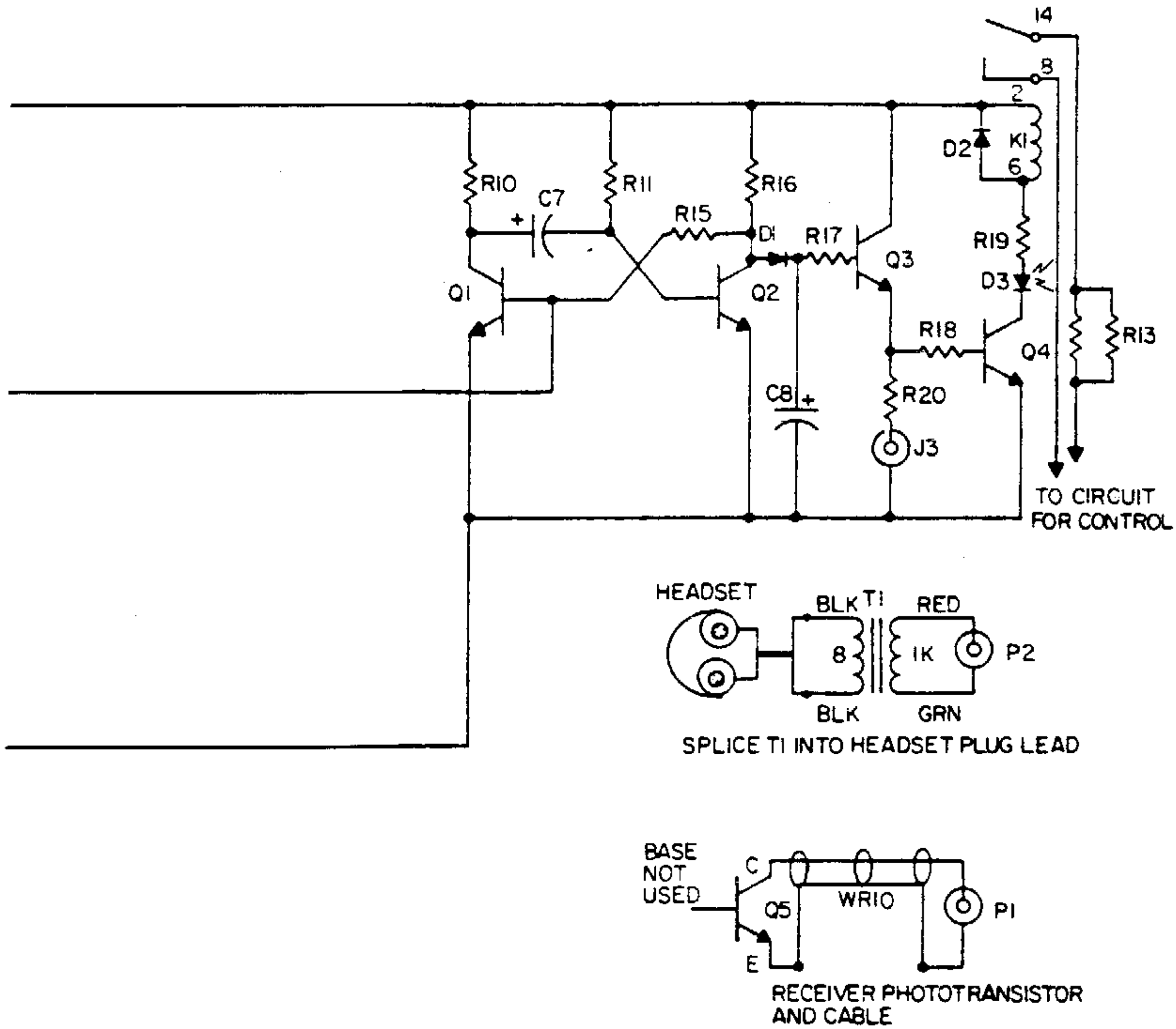
TEST POINT	VALUE
A	6–8V
B	4–5V
C	3–4V
D	0(.3)V
E	9(4.5)V
F	0(3)V
G	0(1)V
H	9(.2)V

VALUE IN PARENTHESIS ARE IN THE "ACTUATED" STATE AND ARE WITH A REALISTIC #CTR-43 RECORDER.

TAB BOOKS INC. *Fig. 50-1*

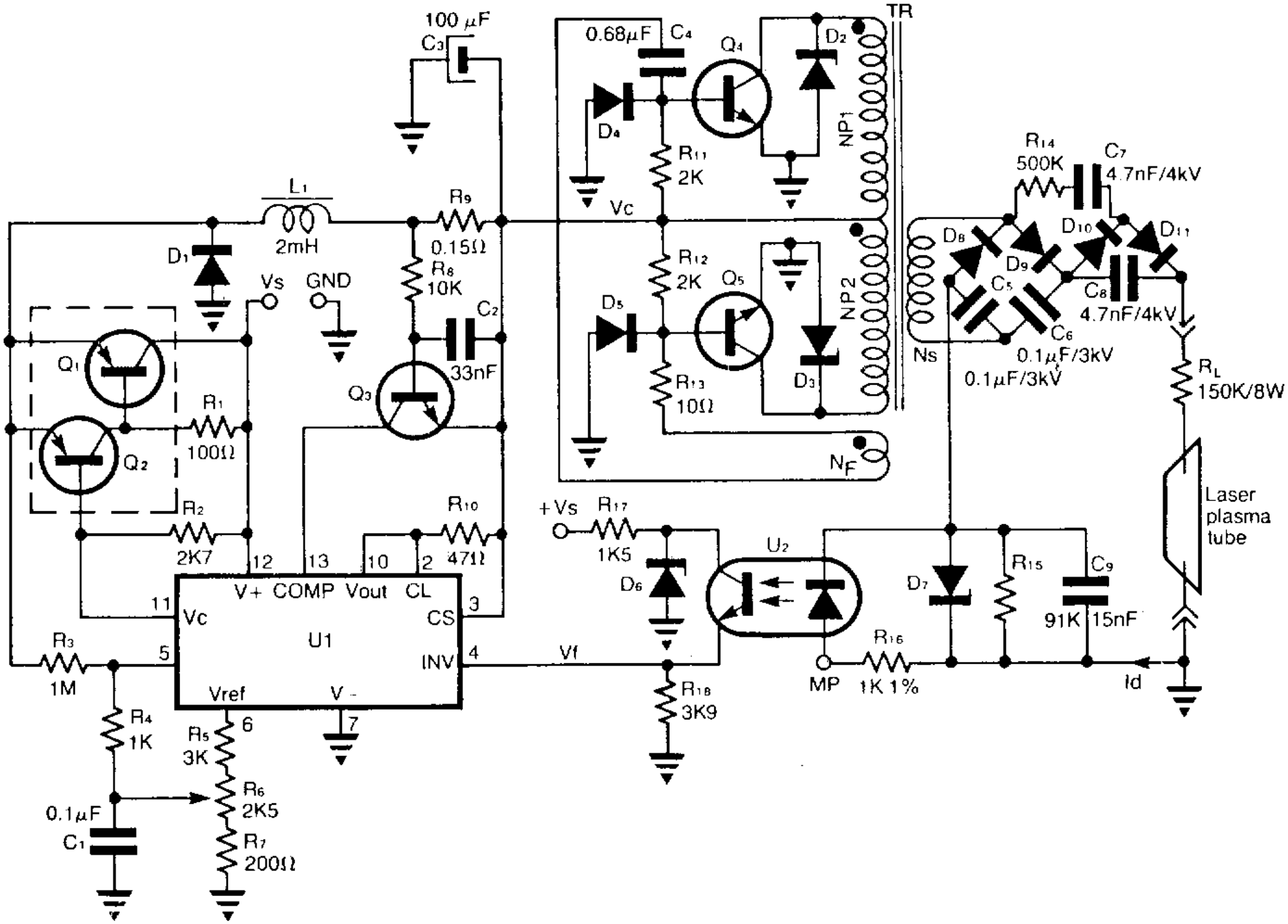
Circuit Notes

The laser light detector utilizes a sensitive photo transistor (Q5) placed at the focal point of a lens (LE2). The output of Q5 is fed to a sensitive amplifier consisting of array (A1) and is biased via the voltage divider consisting of R14 and R1. The base is not used. Q5 is capacitively coupled to a Darlington pair for impedance transforming and is further fed to a capacitively coupled cascaded pair of common-emitter amplifiers for further signal



amplification. Sensitivity control (R7) controls base drive to the final transistor of the array and hence controls overall system sensitivity. Output of the amplifier array is capacitively coupled to a one-shot consisting of Q1 and Q2 in turn integrating the output pulses of Q2 onto capacitor C8 through D1. This dc level now drives relay drivers Q3 and Q4 activating K1 along with energizing indicator D3, consequently controlling the desired external circuitry. The contacts of K1 are in series with low ohm resistor R13 to prevent failure when switching capacitive loads. J2 allows "listening" to the intercepted light beam via headsets. This is especially useful when working with pulsed light sources such as GaAs lasers or any other varying periodic light source.

STABILIZING A LASER DISCHARGE CURRENT



ELECTRONIC ENGINEERING

Fig. 50-2

						D_1	BA159
						D_2; D_3	PL33Z
				Q_1	KD167	D_4; D_5	1N4148
		$Np_1 = Np_2$	8 turns	Q_2	BD140	D_6	PL7V5
U_1	LM723	Ns	1100 turns	Q_3	BC173C	D_7	PL47Z
U_2	MB101	Nf	5 turns	Q_4; Q_5	2N5496	D_8...D_{11}	LA40

Circuit Notes

The circuit uses a free-running push-pull dc to dc high voltage converter to get the necessary voltage for the laser plasma tube supply. The supply voltage V_C of this converter, is adjusted by a switch-mode power supply in order to keep the load current constant, at set value. The linear opto-electronic isolator U2, connected in series with the laser plasma tube, gives a voltage V_F proportional to the discharge current I_D across R18, having the correct polarity to drive directly the inverting input of U1, D7, R15 protects the optoisolator diode against damage produced by the high voltage ignition pulse.

Due to the high operating frequency of the high voltage converter (25 kHz) the ripple of the laser output power is less than 2.10^{-4}. The stability of I_D is better than 10^{-2}, for variations of supply voltage V_s is the range of $\pm 10\%$, and depends on the optoisolator sensitivity.

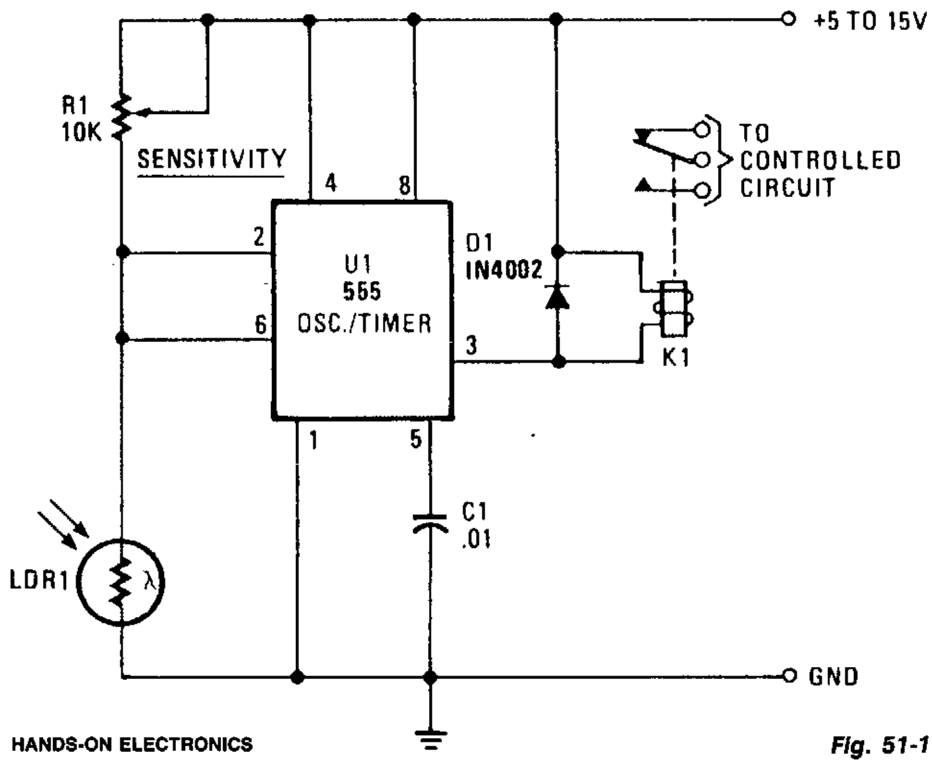
51

Light-Controlled Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Photo Alarm
Warning Light Operates from Battery Power Supply
Light Operated Switch
Photoelectric Switch
Back-Biased GaAsP LED Operates as Light Sensor
Twilight-Triggered Circuit
Automatic Mooring Light
Electronic Wake-Up Call
Photodiode Sensor Amplifier
Light Seeking Robot
Synchronous Photoelectric Switch
Photocurrent Integrator
Robot Eyes
Modulated Light-Beam Circuit Cancels Ambient Light Effects
Monostable Photocell Circuit has Self-Adjusting Trigger Level
Thermally Stabilized PIN Photodiode Signal Conditioner

PHOTO ALARM



HANDS-ON ELECTRONICS

Fig. 51-1

Circuit Notes

LDR1, a cadmium sulphide (CDS) photoresistive cell is used as the lower leg of a voltage divider between V_{CC} and ground. The timer terminals 2 and 6 are connected to the junction of the photocell and SENSITIVITY control R1. The resistance of the photoresistive cell varies inversely as the light intensity; resistance is high when the illumination level is low; low in bright light. (The Radio Shack CDS cell 276-116 has a typically wide resistance range—about 3 megohms in darkness and 100 ohms in bright light.) When the light is interrupted or falls below a level set by SENSITIVITY control R1, the rise in LDR1's resistance causes the voltage on pins 2 and 6 to rise. If the control is set so the voltage rises above ⅔ V_{CC}, the relay pulls in. The relay drops out when the light level increases and the drop across the photocell falls below ⅓ V_{CC}. (The circuit can be modified by placing relay K1 and diode D1 between pin 3 and ground. In this case, the relay drops out when the voltage on pins 2 and 6 rises above ⅔ V_{CC}, and pulls in when it falls below ⅓ V_{CC}. This modification is valuable when the relay has single-throw contacts.) Opening and closing of the relay contacts occurs at different illumination levels. This ⅓ V_{CC} hysteresis is an advantage that prevents the circuit from hunting and the relay from chattering when there are very small changes in illumination.

WARNING LIGHT OPERATES FROM BATTERY POWER SUPPLY

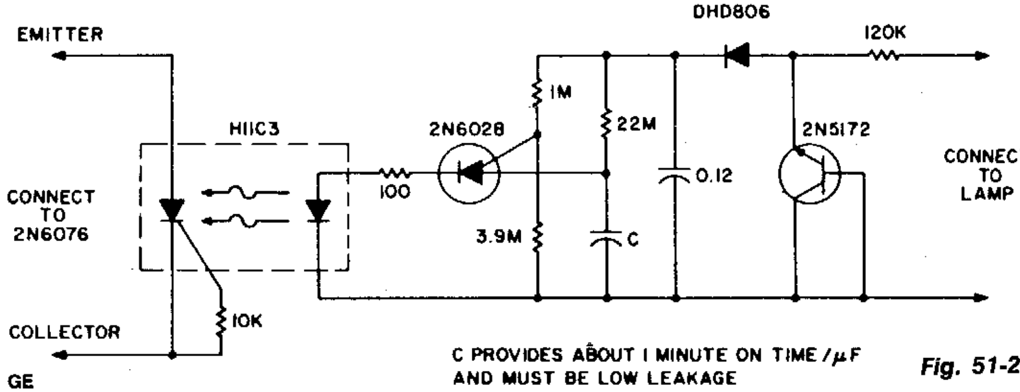
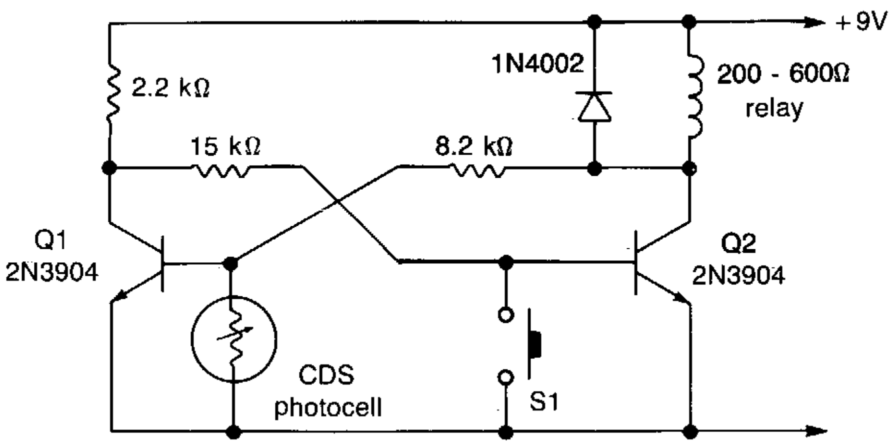


Fig. 51-2

Circuit Notes

The circuit provides illumination when darkness comes. By using the gain available in darlington transistors, this circuit is simplified to use just a photodarlington sensor, a darlington amplifier, and three resistors. The illumination level will be slightly lower than normal, and longer bulb life can be expected, since the D40K saturation voltage lowers the lamp operating voltage slightly.

LIGHT-OPERATED SWITCH



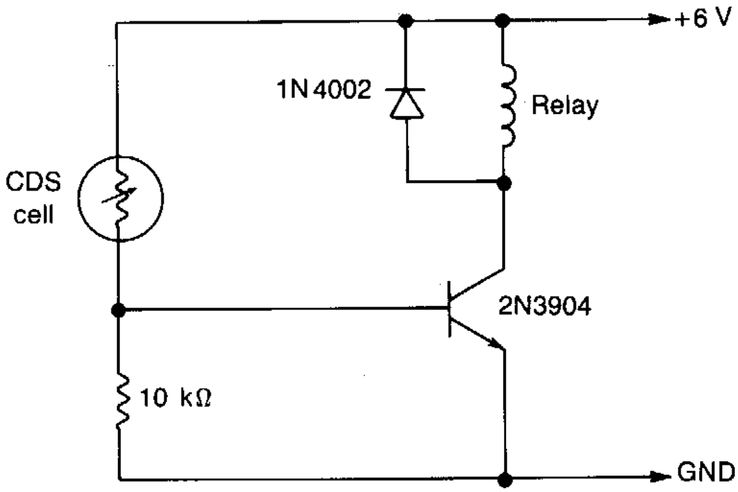
WILLIAM SHEETS

Fig. 51-3

Circuit Notes

This circuit uses a flip-flop arrangement of Q1 and Q2. Normally Q1 is conducting heavily. Light on CDS photocell causes Q1 bias to decrease, cutting it off, turning on Q2, removing the remaining bias from Q1. Reset is accomplished by depressing S1.

PHOTOELECTRIC SWITCH



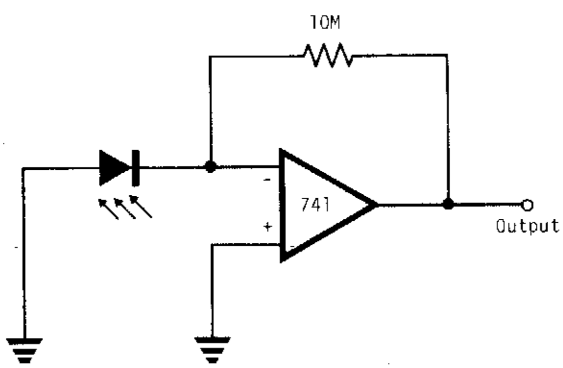
WILLIAM SHEETS

Fig. 51-4

Circuit Notes

The CDS cell resistance decreases in the presence of light, turning on the 2N3904 relay driver.

BACK-BIASED GaAsP LED OPERATES AS LIGHT SENSOR



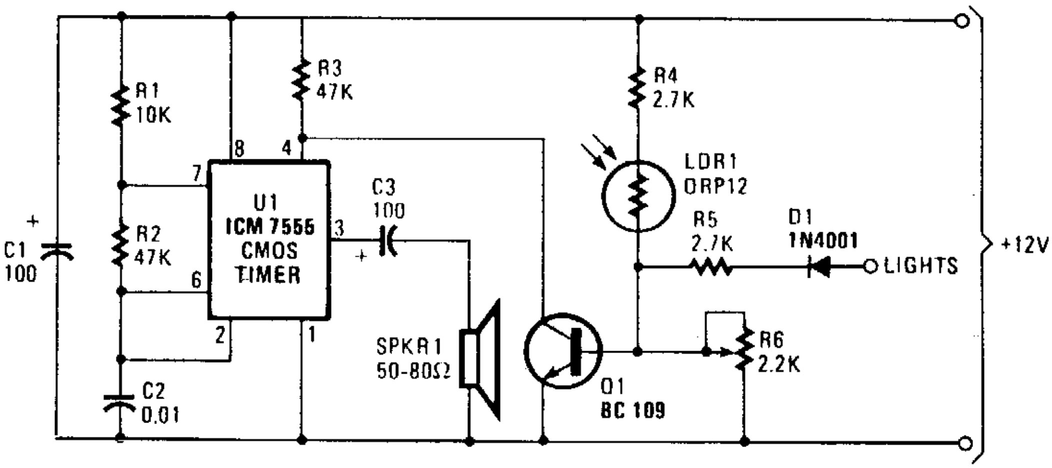
ELECTRONIC ENGINEERING

Fig. 51-5

Circuit Notes

Using a simple 741 amplifier connected as a current-to-voltage converter with the LED as the current source, the voltage at the output is proportional to incident light. The junction is biased only by the difference between the summing node junction potential and ground, preventing the possibility of reverse breakdown. The photon-generated current equals the short-circuit current of the junction, which is linearly related to incident light. The sensor requires a level of incident illumination that depends on the degree of opacity of the LED package.

TWILIGHT-TRIGGERED CIRCUIT



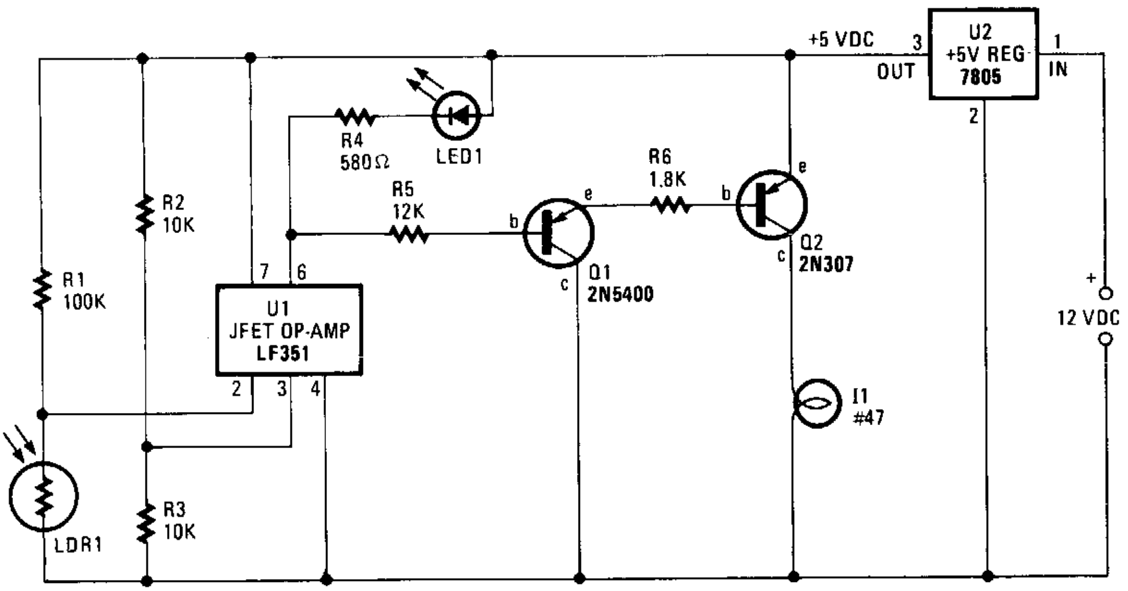
HANDS-ON ELECTRONICS

Fig. 51-6

Circuit Notes

As dusk begins to fall, the sensor (a cadmium-sulfide light-dependent resistor or LDR) operates a small horn to provide an audible reminder that it's time to turn on your lights. To turn the circuit off—simply turn your headlights on and the noise stops. The base of Q1 is fed through a voltage divider formed by R4, LDR1—a light-dependent resistor with an internal resistor of about 100 ohms under bright-light conditions and about 10 megohms in total darkness—potentiometer R6. Q1's base voltage depends on the light level received by LDR1 and the setting of R6. If LDR1 detects a high light level, its resistance decreases, thereby providing a greater base current for Q1, causing it to conduct. When Q1 conducts, pin 4 of U1 is pulled to near ground potential, muting the oscillator. If, on the other hand, LDR1 detects a low light level, its resistance increases (reducing base currentto Q1), cutting off the transistor and enabling the oscillator. In actual practice, you set R6 so that at a suitable light level (dusk), the oscillator will sound. The anode of diode D1 connects to the light switch, where it connects to the vehicle's parking lights. With the lights switched off, that point is connected to the negative chassis by way of the parking lamp. That has no effect on the circuit, as D1 blocks any current flow to ground from Q1's base via R6 and the sidelight lamps. When the lights are switched on, the anode of D1 is connected to the positive supply via the parking lamp switch, thereby applying a voltage to the base of Q1, biasing it into conduction. With Q1 conducting, pin 4 of U1 is pulled virtually to ground, disabling the oscillator even though LDR1's resistance is not enough to do so.

AUTOMATIC MOORING LIGHT



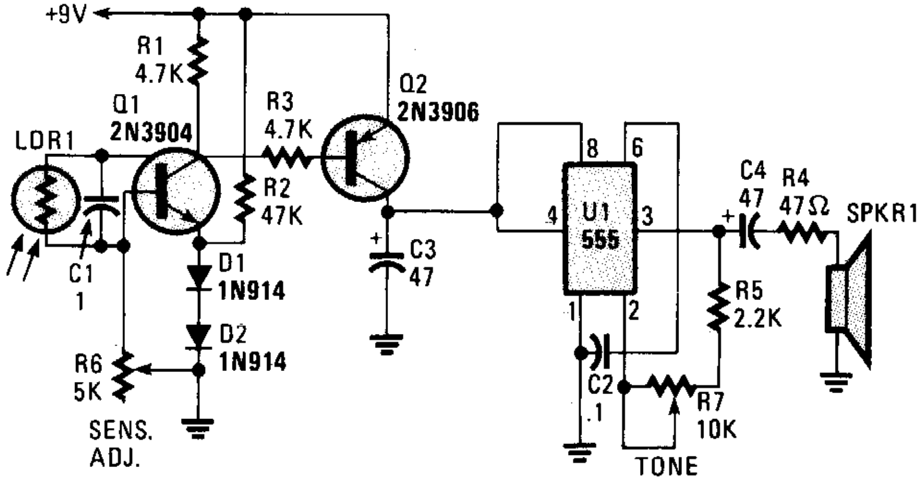
HANDS-ON ELECTRONICS

Fig. 51-7

Circuit Notes

Integrated-circuit U1—an LF351 or 741 op amp—is used as a comparator to control the light. Resistors R2 and R3 provide a reference voltage of about 2.5 volts at pin 3 of U1. When daylight falls on light-dependent resistor LDR1, its resistance is low: about 1000 ohms. In darkness, the LDR's resistance rises to about 1 megohm. Since R1 is 100,000 ohms, and the LDR in daylight is 1000 ohms, the voltage ratio is 100 to 1; the voltage drop across the LDR is less than the 2.5 volt reference voltage and pin 2 of U1 is held at that voltage. In that state, the output at pin 6 of U1 is positive at about 4.5 volts, a value that reverse-biases Q1 to cutoff, which in turn holds Q2 in cutoff, thereby keeping lamp I1 off. When darkness falls, the LDR's resistance rises above R1's value and the voltage at pin 2 of U1 rises above the reference voltage of 2.5 volts. U1's output terminal (pin 6) falls to less than a volt and Q1 is biased on. The base-to-emitter current flow turns Q2 on, which causes current to flow through the lamp. When daylight arrives, the LDR's resistance falls sharply, which causes the lamp to be turned off, ready to repeat the next night/day cycle.

ELECTRONIC WAKE-UP CALL



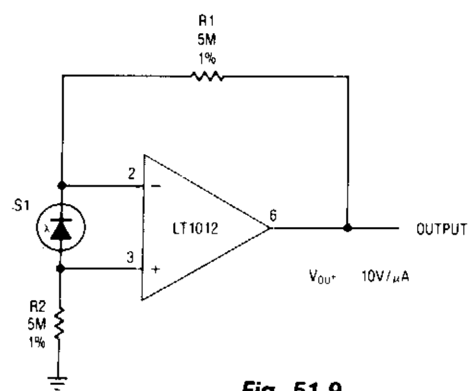
HANDS-ON ELECTRONICS

Fig. 51-8

Circuit Notes

A cadmium sulfide photocell (LDR1, which is a light-dependent resistor) is connected to the base and collector of an npn transistor, Q1. When light hits LDR1, the internal resistance goes from a very high (dark) value to a low (light) value, supplying base current to Q1, turning it on. The voltage across R1 produces a bias that turns Q2 on, which in turn, supplies the positive voltage to U1 at pin 8 (the positive-supply input) and pin 4 (the reset input), to operate the 555 audio oscillator circuit. The circuit's sensitivity to light can be set via R6 (a 50,000 ohm potentiometer). R7 sets the audio tone to the most desirable sound. The squarewave audio tone is fed from U1 pin 3 to a small speaker through coupling capacitor C4 and current limiting resistor R4.

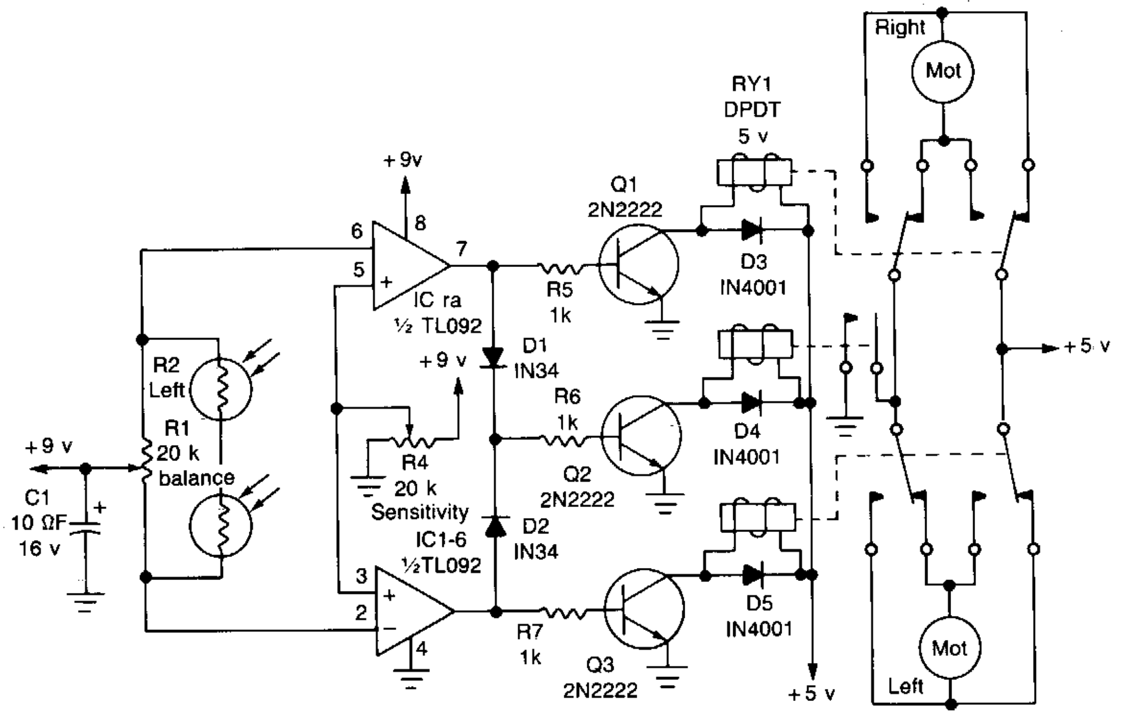
PHOTODIODE SENSOR AMPLIFIER



LINEAR TECHNOLOGY CORP.

Fig. 51-9

LIGHT-SEEKING ROBOT



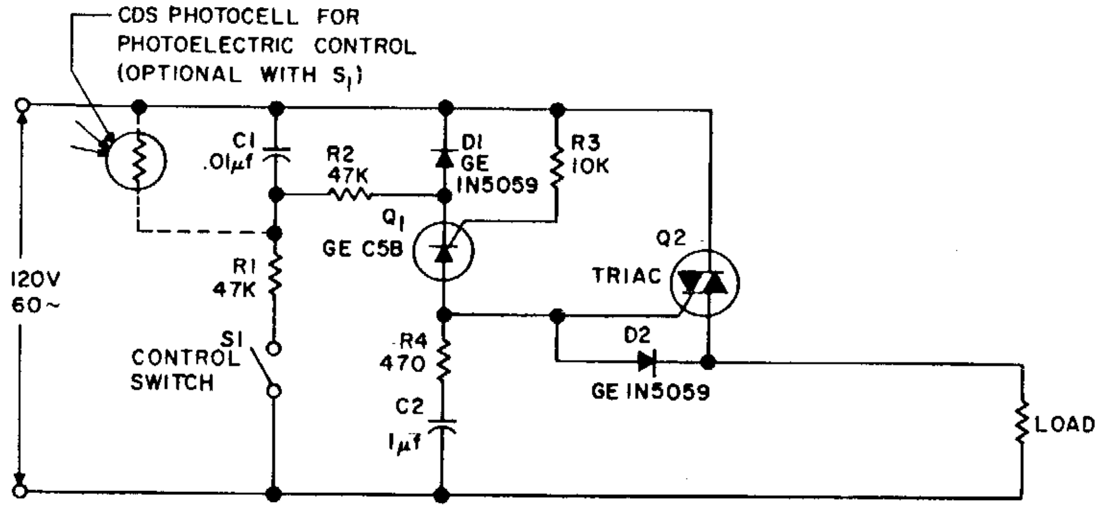
RADIO-ELECTRONICS

Fig. 51-10

Circuit Notes

The circuit is light seeking; it will follow a flashlight around a darkened room. A pair of photocells determine the direction in which the robot will move. Each photocell is connected to an op amp configured as a comparator. When sufficient light falls on photocell R2, the voltage at the inverting input (pin 6) of IC1-a will fall below the voltage at the non-inverting input (pin 5), so the output of the comparator will go high, and transistors Q1 and Q2 will turn on. That will enable relays RY1 and RY2, and thereby provide power for the right motor. The robot will then turn left. Likewise, when light falling on R3 lowers its resistance, Q2 and Q3 will turn on, the left motor will energize, and the robot will turn right.

SYNCHRONOUS PHOTOELECTRIC SWITCH



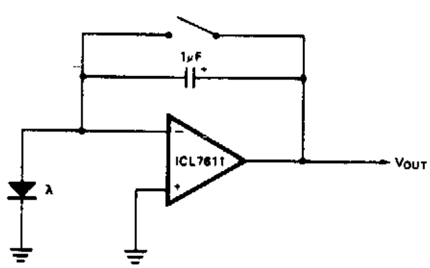
GENERAL ELECTRIC

Fig. 51-11

Circuit Notes

Synchronous switching is turning on only at the instant the ac supply voltage passes through zero, and turning off only when current passes through zero. This circuit provides this function in response to either a mechanical switch or a variable resistance such as a cadmium-sulfide photocell. This circuit produces the minimum disturbance to the power supply when switching, and always conducts an integral number of whole cycles. It is ideal for use wherever RFI and audio filtering is undesirable, where magnetizing inrush current of transformers causes nuisance fuse-blowing, and where sensitive equipment must operate in the vicinity of power switches.

PHOTOCURRENT INTEGRATOR



0307-32

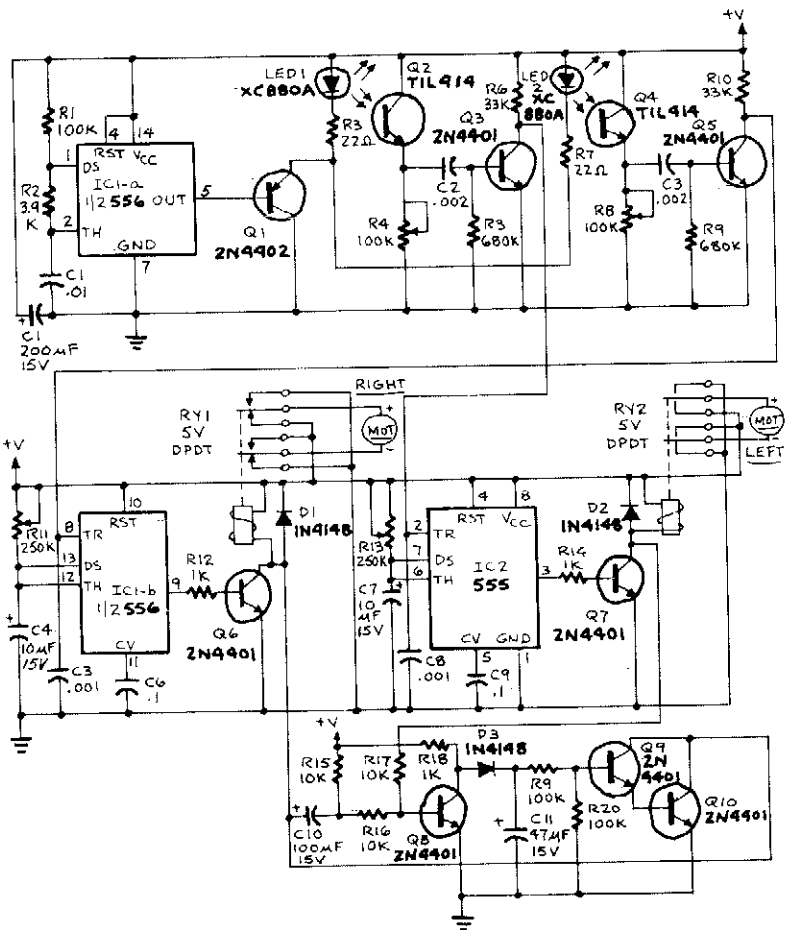
Circuit Notes

Low leakage currents allow integration times up to several hours.

INTERSIL

Fig. 51-12

ROBOT EYES



RADIO ELECTRONICS

Fig. 51-13

Circuit Notes

An infrared LED and a phototransistor are used for each eye. Half of a 556 timer IC (IC1-a) functions as an astable multivibrator oscillating at a frequency of about 1 kHz. That IC drives transistor Q1 which in turn drives the two infrared LED's, LED1 and LED2. The right eye is composed of LED1 and Q2. If an obstacle appears in front of the right eye, pulses from LED1 are reflected by the obstacle and detected by Q2. The signal from Q2 is amplified by Q3, which triggers IC2, a 555. That IC operates in the monostable mode, and it provides a pulse output with a width of as much as 2.75 seconds, depending on the setting of R11. That pulse output energizes relay RY1, and that reverses the polarity of the voltage applied to the motor. Corresponding portions of the circuit of the left eye operate in the same fashion, using the unused half of the 556 (IC1-b). That action causes the robot to turn away from an obstacle.

MODULATED LIGHT-BEAM CIRCUIT CANCELS AMBIENT LIGHT EFFECTS

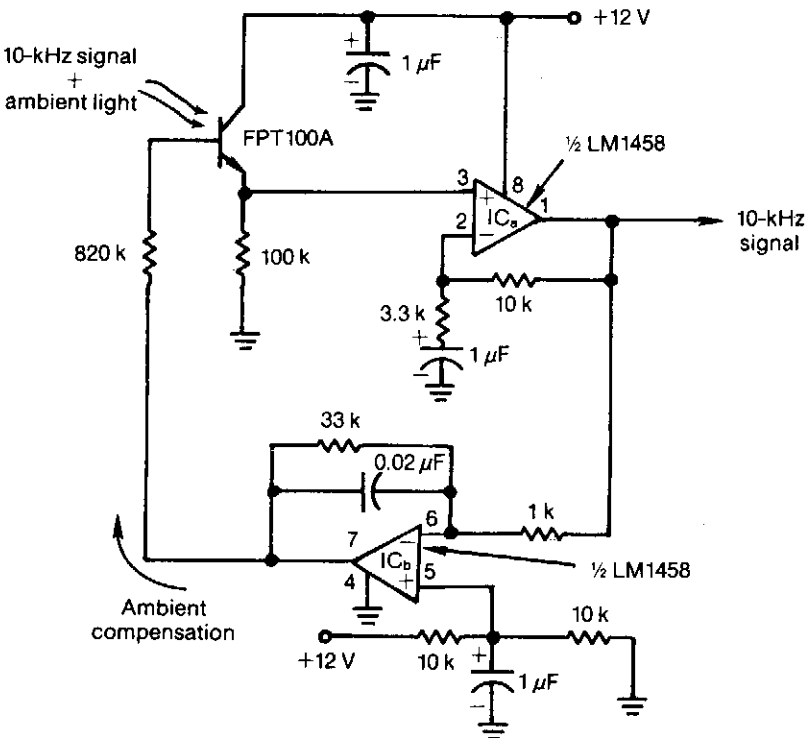


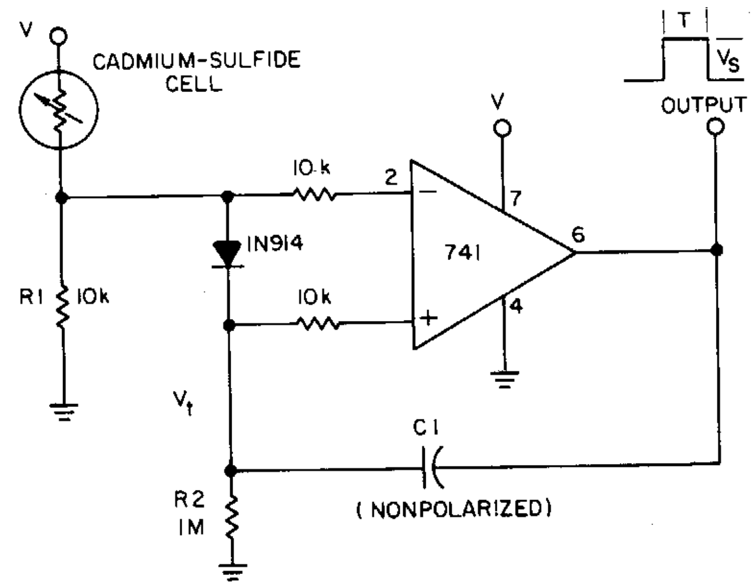
Fig. 51-14

ELECTRONIC DESIGN

Circuit Notes

Feedback control of the phototransistor in this optical detector helps negate the effects of varying ambient light sources. The output of a modulated visible-light LED is detected, amplified, buffered, and fed through a low-pass filter. Ambient light signals below the LED's 10-kHz modulating rate reach the detector's base out of phase with incoming ambient light and cancel the undesired effects.

MONOSTABLE PHOTOCELL CIRCUIT HAS SELF-ADJUSTING TRIGGER LEVEL



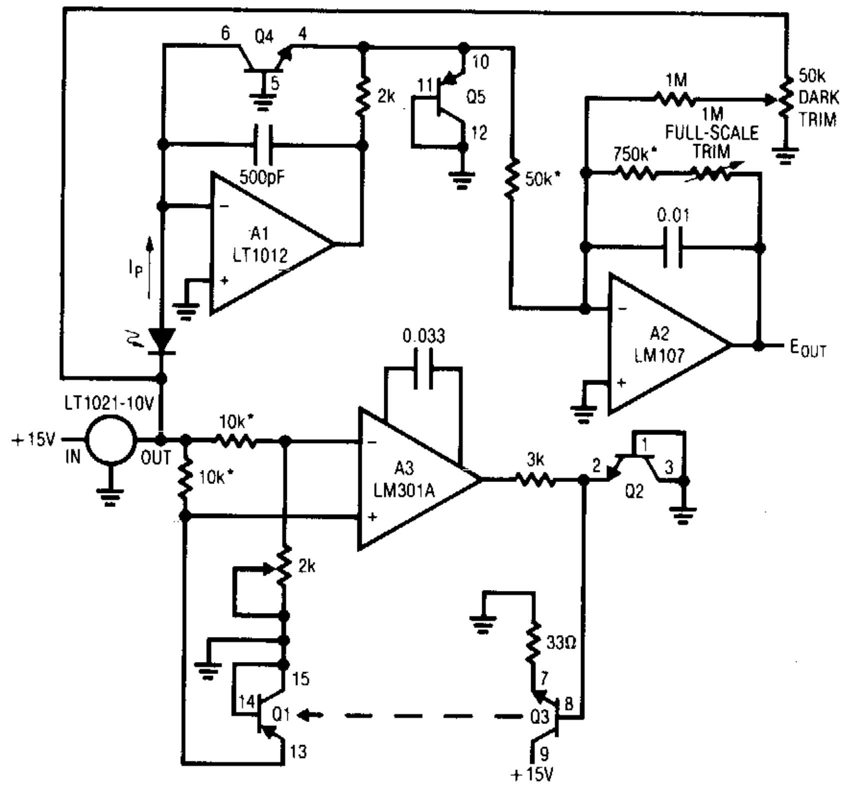
ELECTRONIC DESIGN

Fig. 51-15

Circuit Notes

A photocell circuit provides automatic threshold adjustment. Monostable action prevents undesired retriggering of the output. With only one op amp IC, the circuit offers: Automatic adjustment of its trigger level to accommodate various light sources, changes in ambient light and misalignments; A built-in monostable action to provide only a single output pulse during a preset time; Feedback action to raise the threshold level after triggering and to speed switching. The feedback also eliminates the circuit's tendency to oscillate during switching.

THERMALLY STABILIZED PIN PHOTODIODE SIGNAL CONDITIONER



LINEAR TECHNOLOGY CORPORATION *Fig. 51-16*

Circuit Notes

The photodiode specified responds linearly to light intensity over a 100 dB range. Digitizing the diodes linearly amplified output would require an A-D converter with 17 bits of range. This requirement can be eliminated by logarithmically compressing the diode's output in the signal conditioning circuitry. A1 and Q4 convert the diode's photocurrent to voltage output with a logarithmic transfer function. A2 provides offsetting and additional gain. A3 and its associated components form a temperature control loop

RESPONSE DATA

LIGHT (900μM)	DIODE CURRENT	CIRCUIT OUTPUT
1mW	350μA	10.0V
100μW	35μA	7.85V
10μW	3.5μA	5.70V
1μW	350nA	3.55V
100nW	35nA	1.40V
10nW	3.5nA	−0.75V

 = HP-5082-4204 PIN PHOTODIODE.

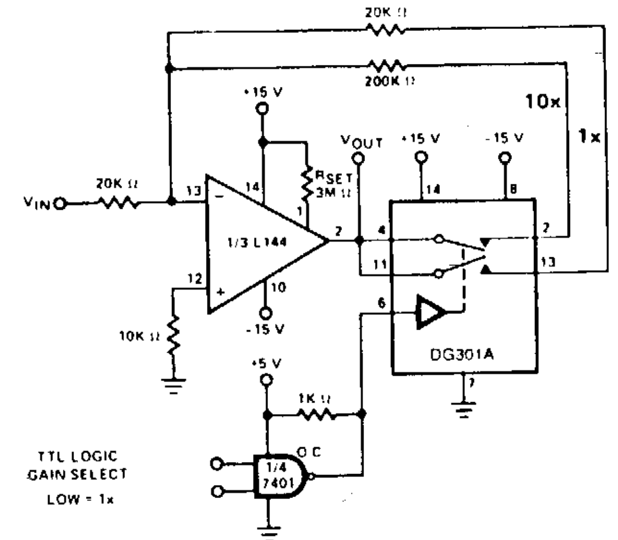
Q1–Q5 = CA3096.
CONNECT SUBSTRATE OF CA3096
ARRAY TO Q4'S EMITTER.
*1% RESISTOR

which maintains Q4 at constant temperature (all transistors in this circuit are part of a CA3096 monolithic array). The 0.033 μF value at A3's compensation pins gives good loop damping if the circuit is built using the array's transistors in the location shown. Because of the array die's small size, response is quick and clean. A full-scale step requires only 250 ms to settle to final value. To use this circuit, first set the thermal control loop. To do this, ground Q3's base and set the 2 k pot so A3's negative input voltage is 55 mV above its positive input. This places the servo's setpoint at about 50°C (25°C ambient + (2.2 mV/°C × 25°C rise = 55 mV = 50°C). Unground Q3's base and the array will come to temperature. Next, place the photodiode in a completely dark environment and adjust the "dark trim" so A2's output is 0 V. Finally, apply or electrically simulate 1 mW of light and set the "full-scale" trim for 10 V out. Once adjusted, this circuit responds logarithmically to light inputs from 10nW to 1mW with an accuracy limited by the diode's 1% error.

52
Logic Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

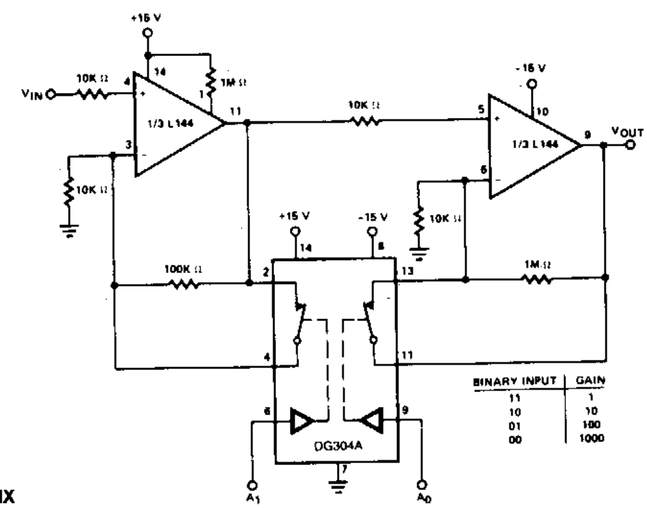
LOW POWER INVERTING AMPLIFIER WITH DIGITALLY SELECTABLE GAIN



SILICONIX

Fig. 52-1

LOW POWER BINARY TO 10^n GAIN LOW FREQUENCY AMPLIFIER



BINARY INPUT	GAIN
11	1
10	10
01	100
00	1000

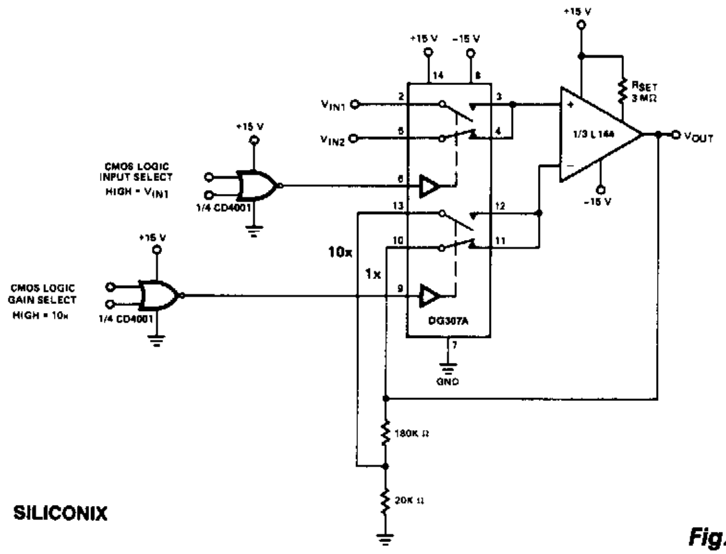
SILICONIX

Fig. 52-2

Circuit Notes

Gain increases by decades as the binary input decreases from 1,1 to 0,0. Minimum gain is 1 and maximum gain is 1000. Since the switch is static in this type of amplifier the power dissipation of the switch will be less than a tenth of a milliwatt.

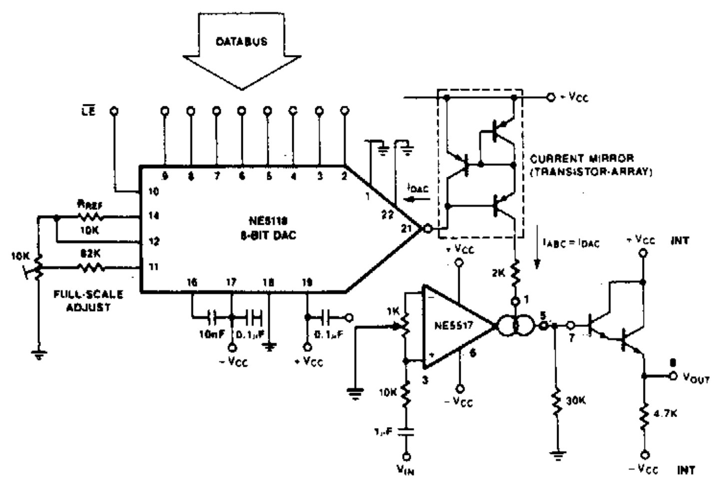
LOW POWER NON-INVERTING AMPLIFIER WITH DIGITALLY SELECTABLE INPUTS AND GAIN



SILICONIX

Fig. 52-3

PROGRAMMABLE AMPLIFIER



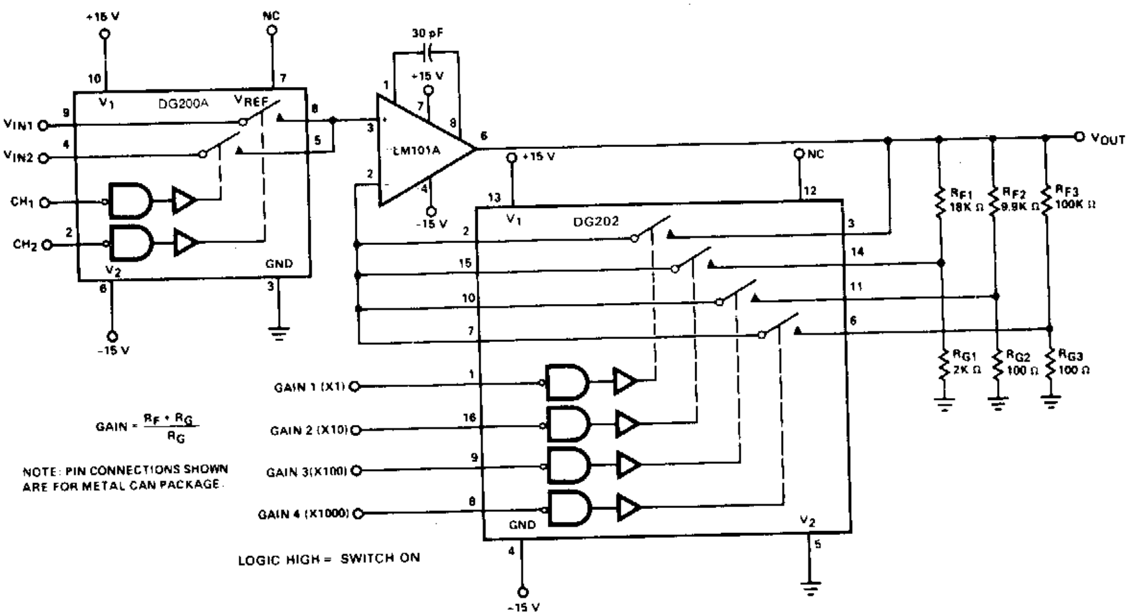
SIGNETICS

Fig. 52-4

Circuit Notes

The intention of the following application shows how the NE5517 works in connection with a DAC. In the application, the NE5118 is used—an 8-bit DAC with current output—its input register making this device fully μP-compatible. The circuit consists of three functional blocks; the NE5118 which generates a control current equivalent to the applied data byte, a current mirror, and the NE5517.

A PRECISION AMPLIFIER WITH DIGITALLY PROGRAMMABLE INPUTS AND GAIN



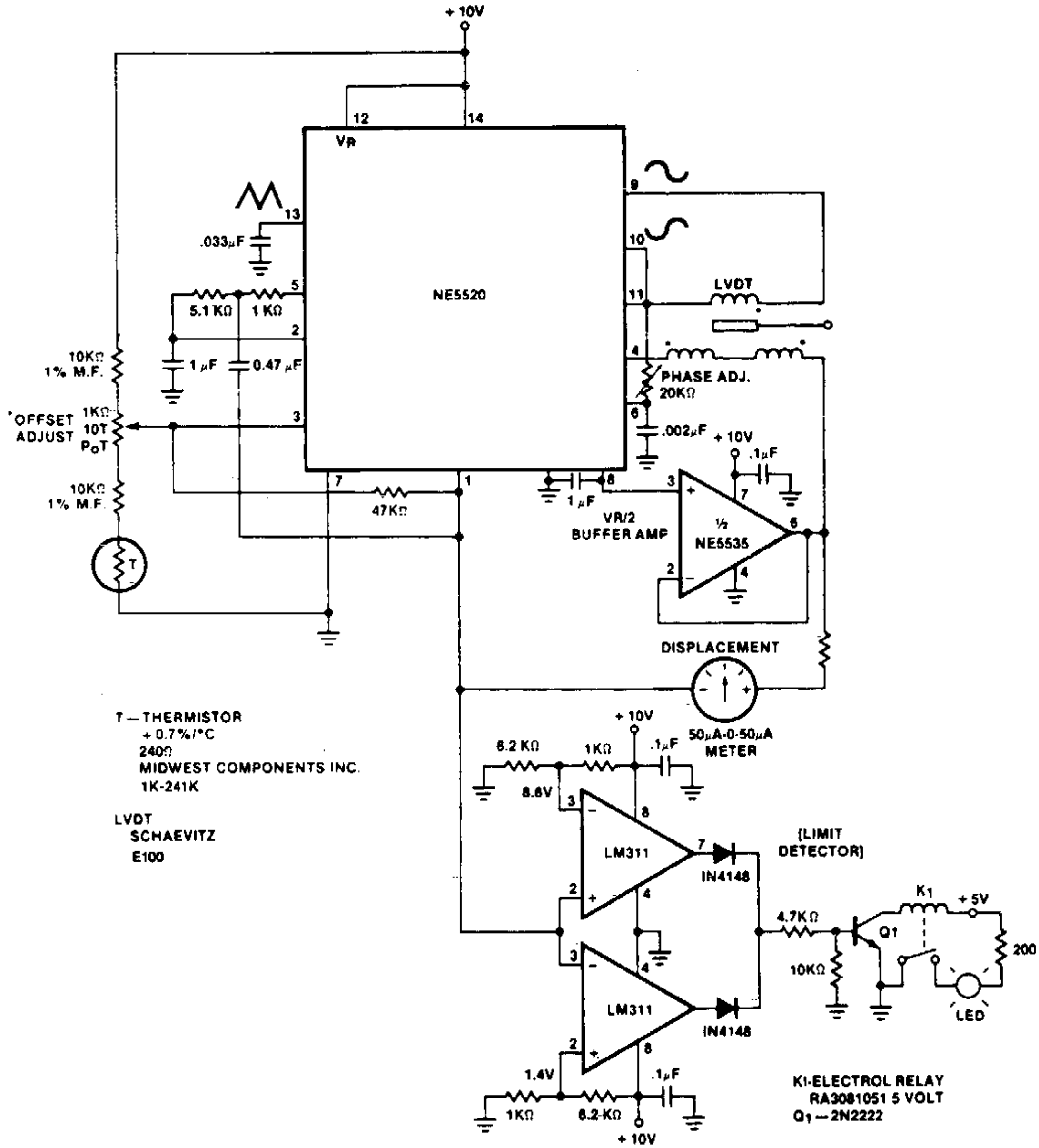
SILICONIX

Fig. 52-5

53
LVDT Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LVDT DRIVER DEMODULATOR



T — THERMISTOR
+0.7%/°C
240Ω
MIDWEST COMPONENTS INC.
1K-241K

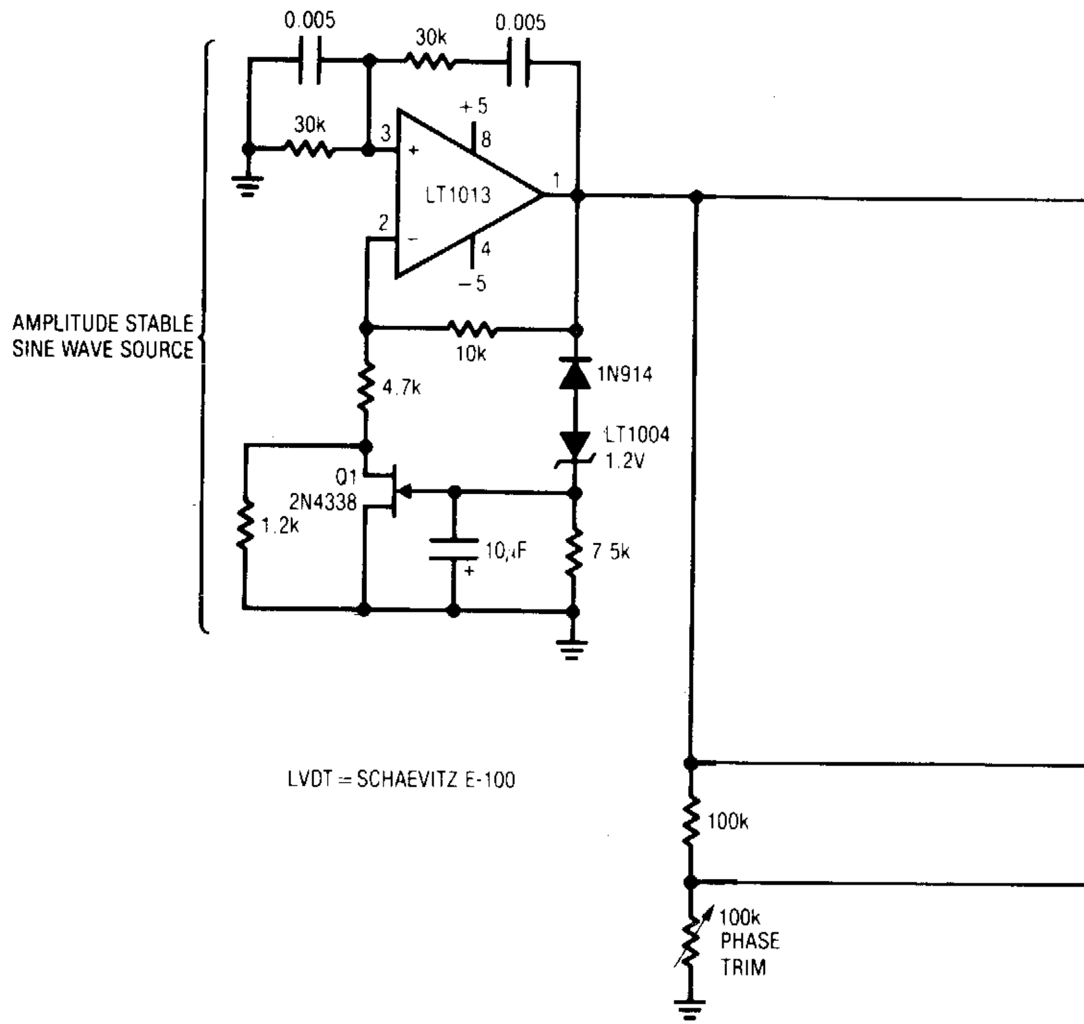
LVDT
SCHAEVITZ
E100

KI-ELECTROL RELAY
RA3081051 5 VOLT
Q1 — 2N2222

SIGNETICS

Fig. 53-1

LINEAR VARIABLE DIFFERENTIAL TRANSFORMER SIGNAL CONDITIONER

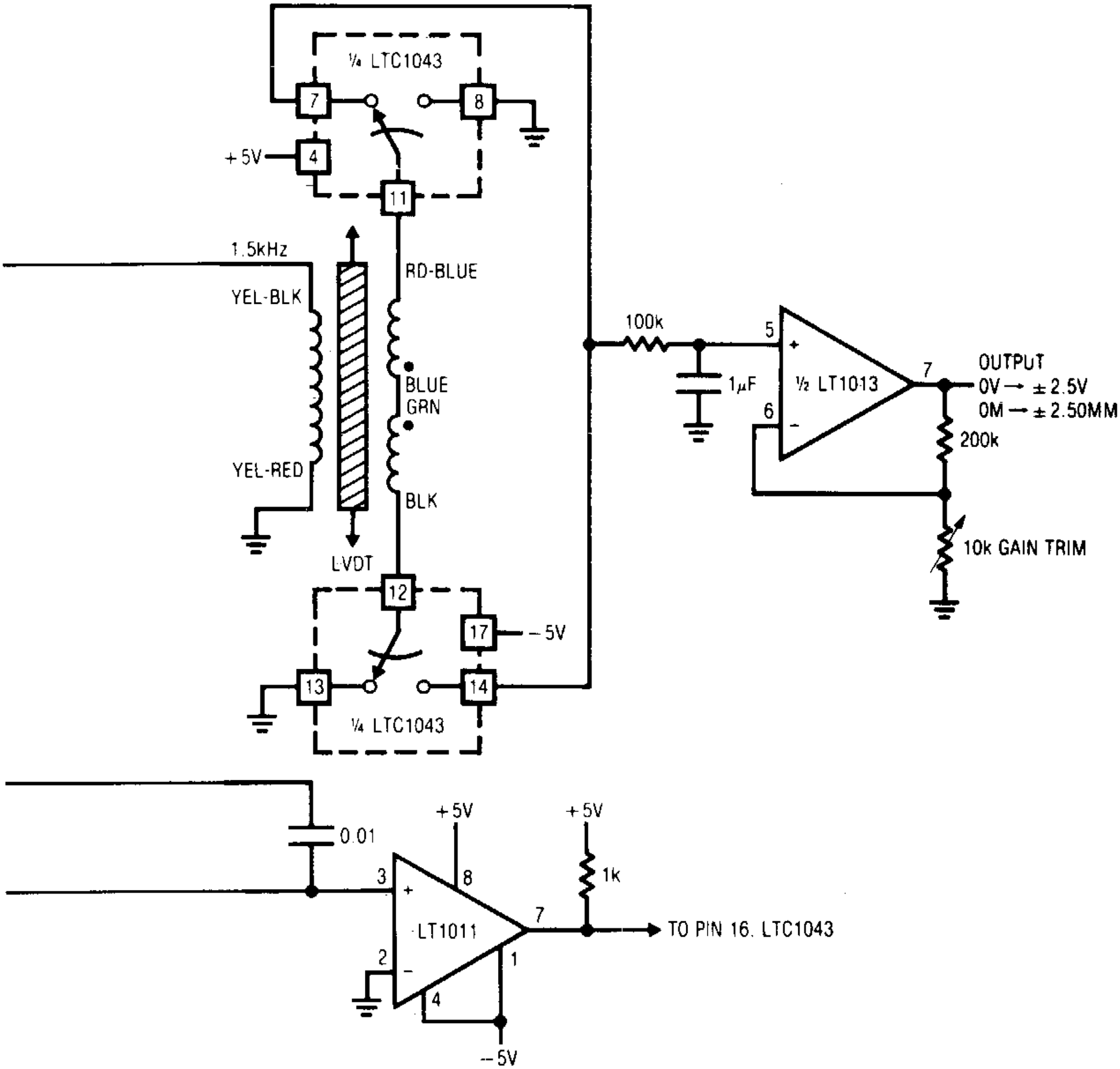


LINEAR TECHNOLOGY CORPORATION

Fig. 53-2

Circuit Notes

A1 and its associated components furnish an amplitude stable sine wave source. A1's positive feedback path is a Wein bridge, tuned for 1.5 kHz, Q1, the LT1004 reference, and additional components in A1's negative loop unity-gain stabilize the amplifier. A1's output an amplitude stable sine wave, drives the LVDT. C1 detects zero crossings and feeds the LTC1043 clock pin. A speed-up network at C1's input compensates LVDT phase shift, synchronizing the LTC1043's clock to the transformer's output zero

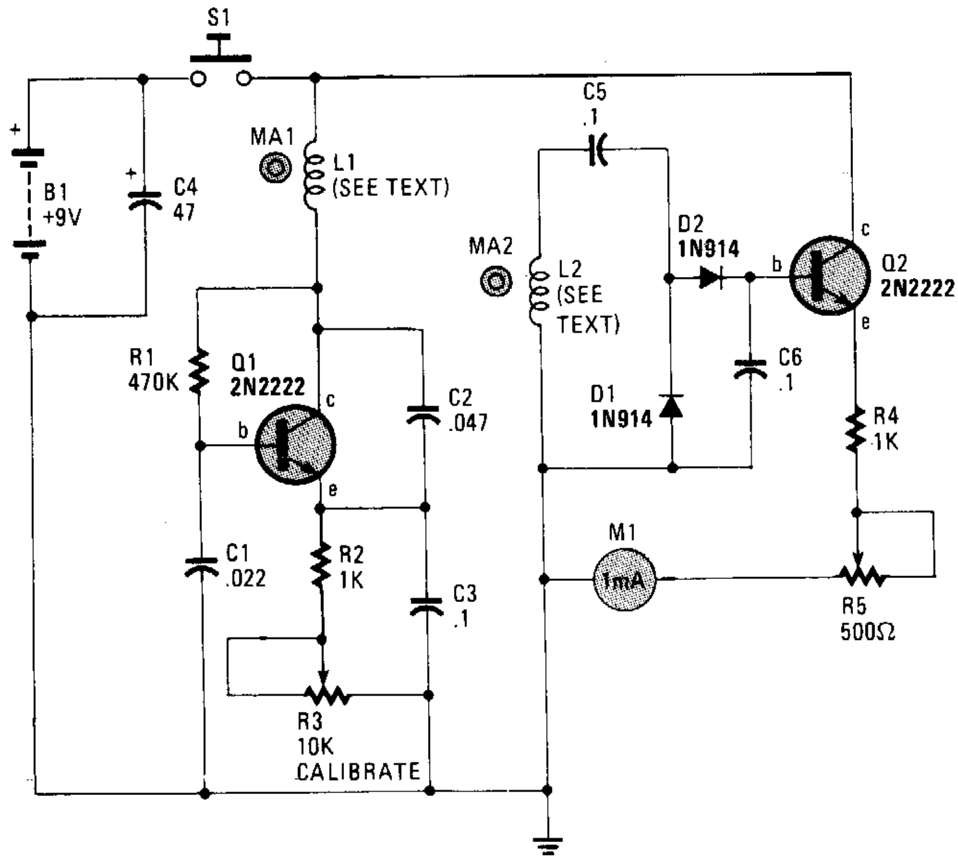


crossings. The LTC1043 alternately connects each end of the transformer to ground, resulting in positive half-wave rectification at pins 7 and 14. These points are summed at a low-pass filter which feeds A2. A2 furnishes gain scaling and the circuit's output. The LTC1043's synchronized clocking means the information presented to the low-pass filter is amplitude and phase sensitive. The circuit output indicates how far the core is from center and on which side. To calibrate this circuit, center the LVDT core in the transformer and adjust the phase trim for 0 V output. Next, move the core to either extreme position and set the gain trim for 2.50 V output.

54
Measuring and Test Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

MAGNETOMETER



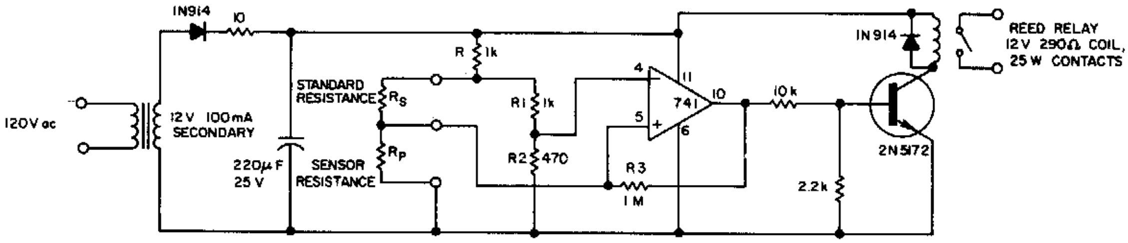
HANDS-ON ELECTRONICS

Fig. 54-1

Circuit Notes

The circuit uses two general-purpose npn transistors, Q1 and Q2, and a special hand-wound, dual-coil probe ferrets out the magnetism. Q1 and its associated components form a simple VLF oscillator circuit, with L1, C2, and C3 setting the frequency. The VLF signal received by the pickup coil, L2, is passed through C5 and rectified by diodes D1 and D2. The small dc signal output from the rectifier is fed to the base of Q2 (configured as an emitter follower), which is then fed to a 0-1 mA meter, M1.

RESISTANCE-RATIO DETECTOR



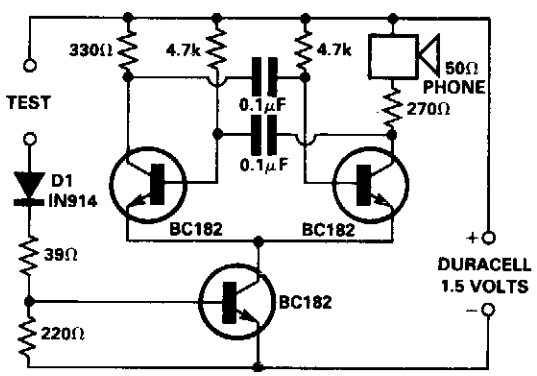
ELECTRONIC DESIGN

Fig. 54-2

Circuit Notes

Applications such as photoelectric control, temperature detection and moisture sensing require a circuit that can accurately detect a given resistance ratio. A simple technique that uses an op amp as a sensing element can provide 0.5% accuracy with low parts cost. The reed-relay contacts close when the resistance of the sensor Rp equals 47% of the standard Rs. Adjusting either R1 or R2 provides a variable threshold; the threshold is controlled by varying R3. For the most part, the type of resistors used for R1 and R2 determines the accuracy and stability of the circuit. With metal-film resistors, less than 0.5% change in ratio sensing occurs over the commercial temperature range (0 to 70 C) with ac input variations from 105 to 135 V.

CONTINUITY TESTER FOR PCB'S



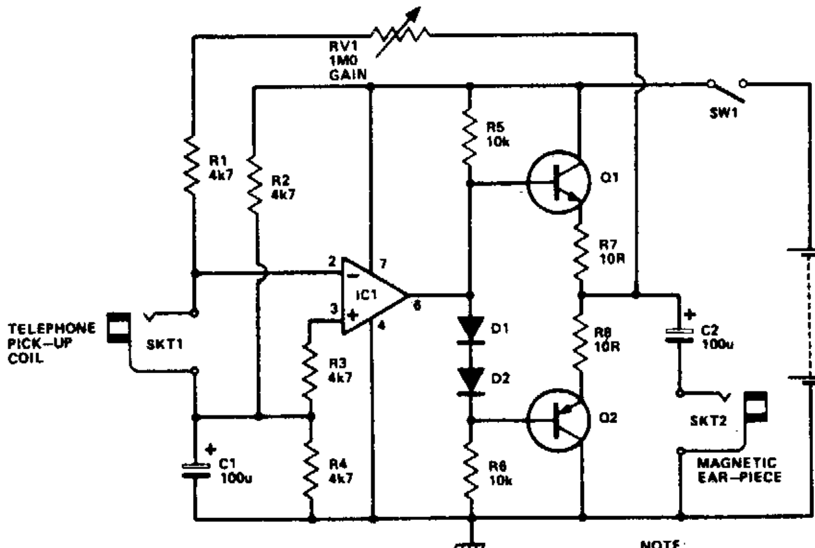
ELECTRONIC ENGINEERING

Fig. 54-3

Circuit Notes

The continuity tester is for tracing wiring on printed circuit boards. It only consumes any appreciable power when the test leads are shorted, so no On/Off switch is used or required. The applied voltage at the test terminals is insufficient to turn on diodes or other semiconductors. Resistors below 50 ohms act as short circuit; above 100 ohms as open circuit. The circuit is a simple multivibrator—T1 and T2, which are switched on by transistor T3. The components in the base of T3 are D1, R1, R2, and the test resistance. With a 1.5 volt supply, there is insufficient voltage to turn on a semiconductor connected to the test terminals. The phone is a telephone earpiece but a 30 ohm speaker would work equally as well.

WIRE TRACER



ELECTRONICS TODAY INTERNATIONAL

Fig. 54-4

Circuit Notes

The tracer detects the weak magnetic field of any current-carrying house wiring and amplifies this signal to a level that is adequate for driving a magnetic earpiece. The unit uses a telephone pick-up coil to detect the magnetic field.

DIODE TESTING

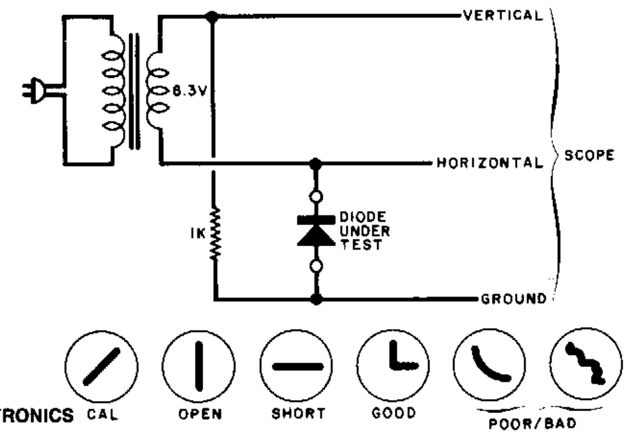


Fig. 54-5

POPULAR ELECTRONICS

Circuit Notes

The circuit will display curves on a scope, contingent on the state of the diode. To "calibrate," substitute a 1000-ohm resistor for the diode and adjust the scope gains for a 45-degree line. The drawings show some expected results.

MEASURING PHASE DIFFERENCE FROM 0° to ±180°

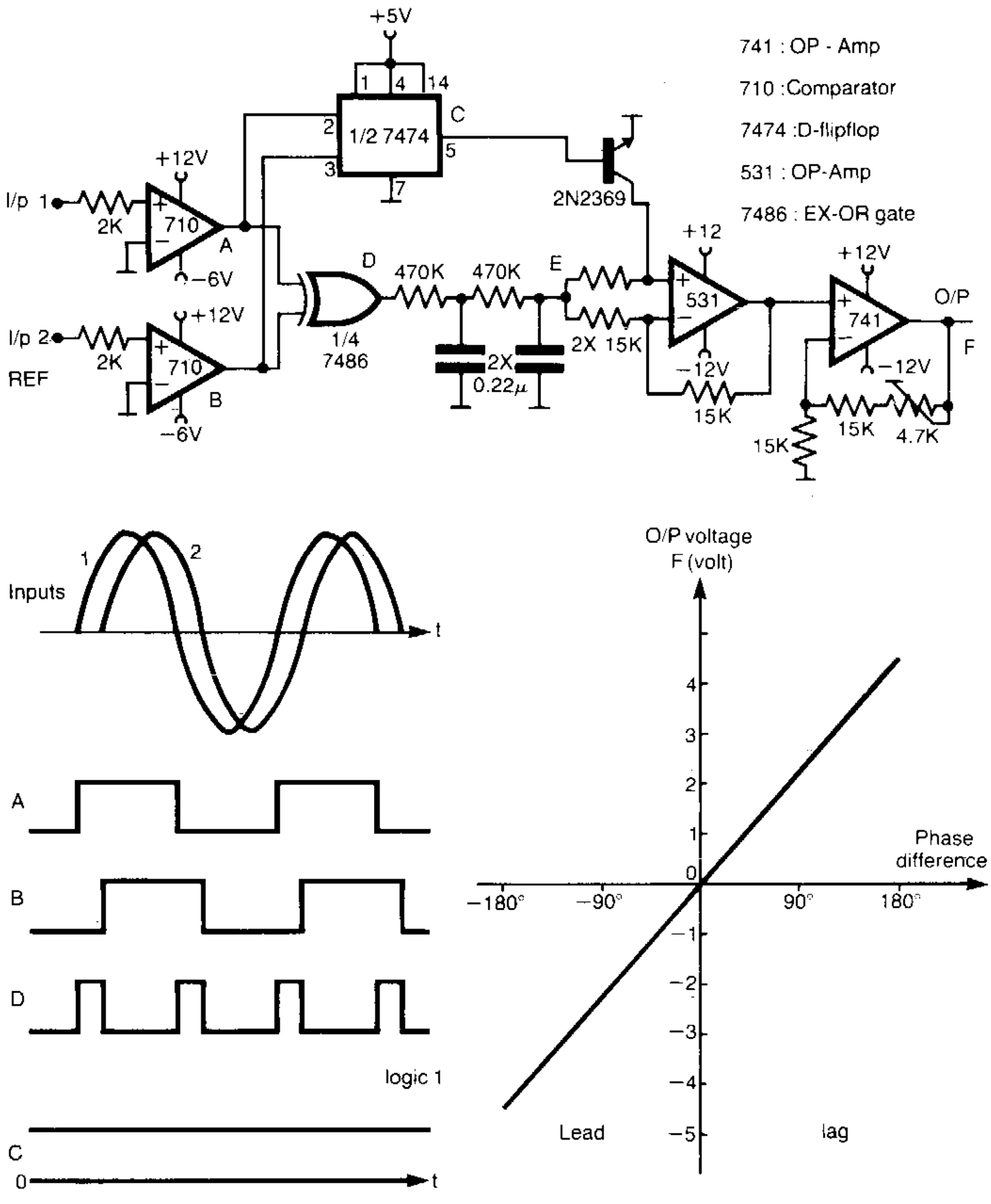


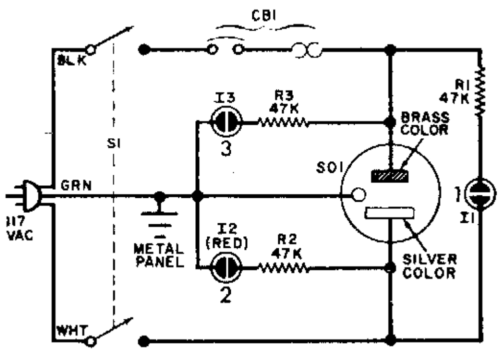
Fig. 54-6

MEASURING PHASE DIFFERENCE FROM 0° to ±180°, Continued.

Circuit Notes

This method is capable of measuring phase between 0 to ±180°. The generated square waves A and B are fed to a D flip-flop which gives an output C equal to logic 1 when input 1 leads input 2 and equal to logic 0 in case of lagging. When C = logic 0, the output of the amplifier F will be positive proportional to the average value E of the output of the EX-OR. When C = logic 1, F will be negative and also proportional to E by the same factor. Hence, the output of the meter is positive in case of lagging and negative for leading. The circuit is tested for sinusoidal inputs and indicates a linearity within 1%. Measurements are unaffected by the frequency of the inputs up to 75 kHz.

GROUND TESTER



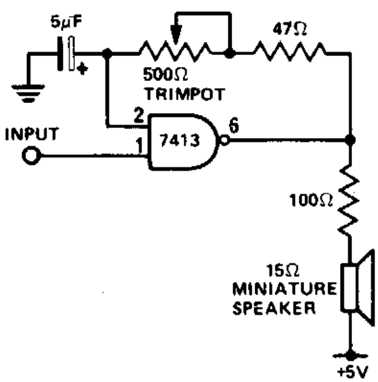
POPULAR ELECTRONICS

Fig. 54-7

Circuit Notes

The circuit is simple and foolproof if wired correctly. Under normal conditions, only lamps 1 and 3 should be lit. If lamp 2 comes on, the cold lead is 117 volts above ground.

MAKING SLOW LOGIC PULSES AUDIBLE



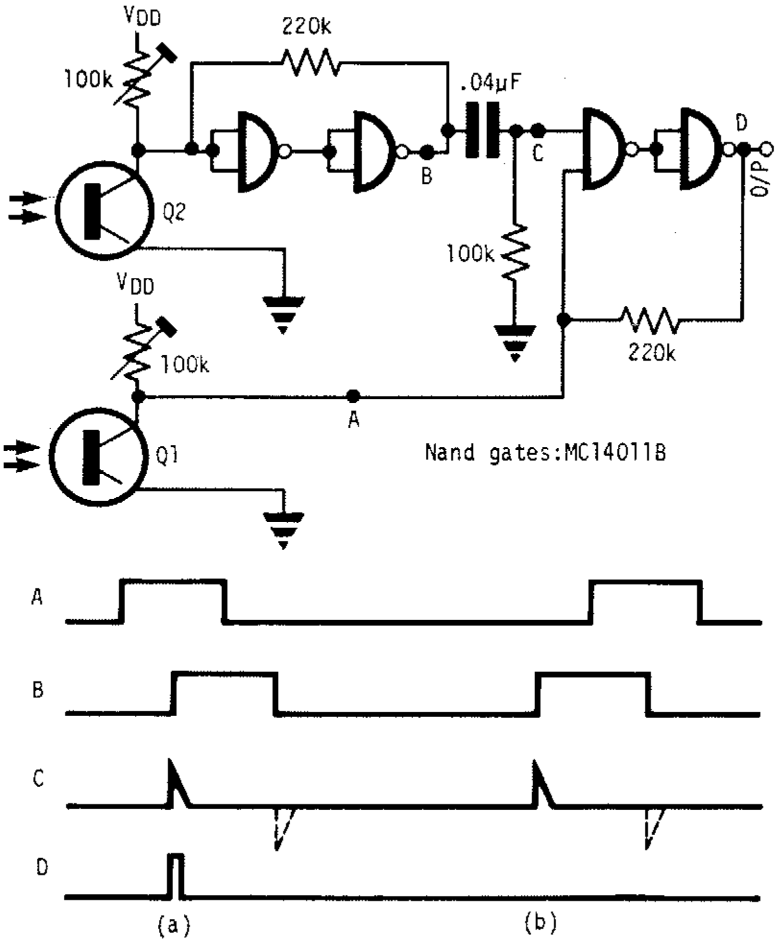
ELECTRONICS TODAY INTERNATIONAL

Fig. 54-8

Circuit Notes

This circuit is useful for monitoring slow logic pulses as a keying monitor or digital clock alarm. The Schmitt trigger is connected as an oscillator. The trimpot controls the pitch of the output. When the input goes high, the circuit will oscillate.

UNIDIRECTIONAL MOTION SENSOR



ELECTRONIC ENGINEERING

Fig. 54-9

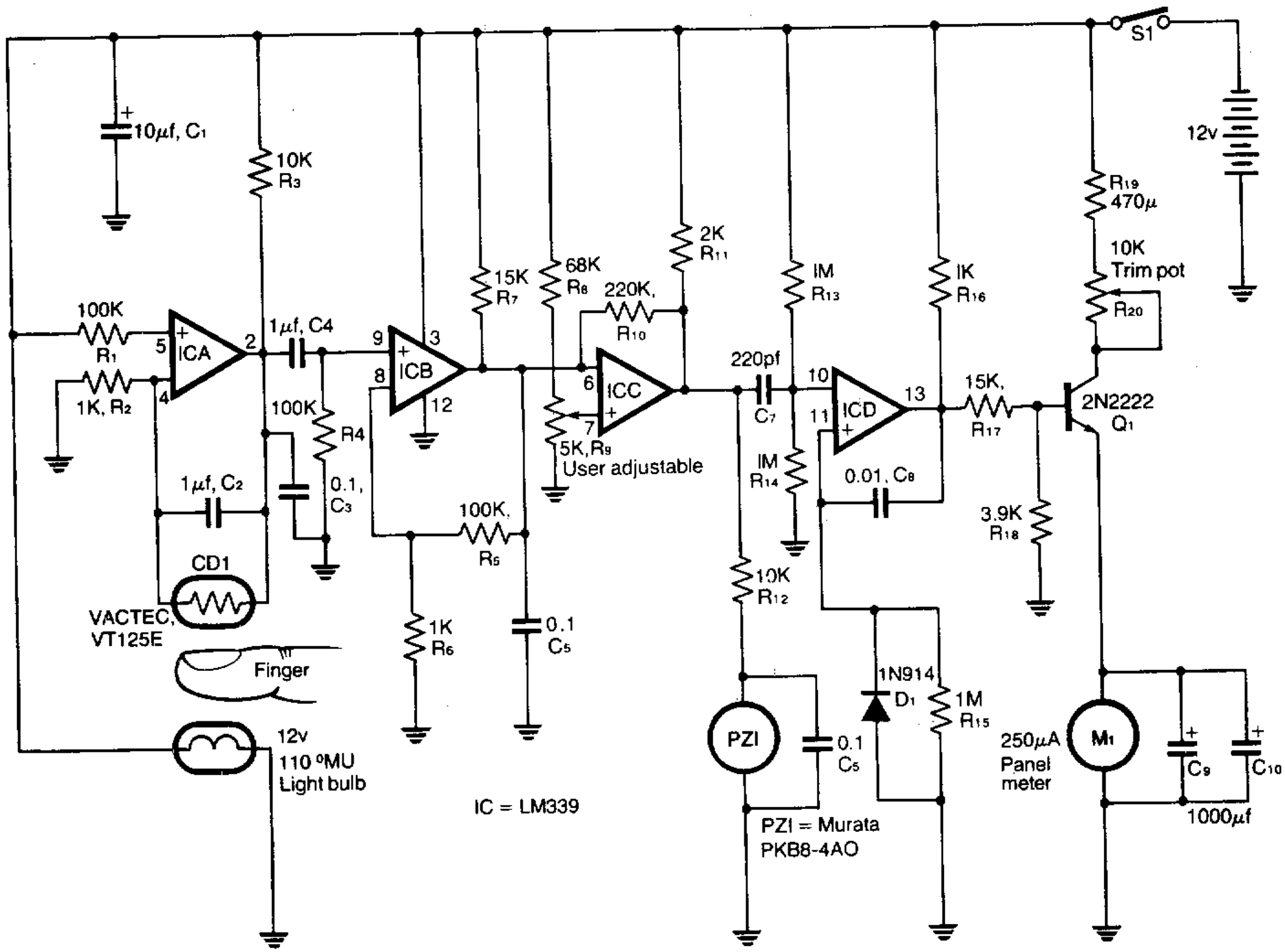
Circuit Notes

This circuit detects an object passing in one direction but ignores it going the opposite way. Two sensors define the sense of direction. The object blocks the light to phototransistor Q1 or Q2 first dependent on the direction of approach. When the object passes Q1 then Q2, an output pulse is generated at D; while no pulse is seen at D as the object passes Q2 then Q1. Object length (measured along the direction of the two sensors) should be greater than the separation of the two sensors Q1 and Q2.

55
Medical Electronics Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

HEART RATE MONITOR



ELECTRONIC ENGINEERING

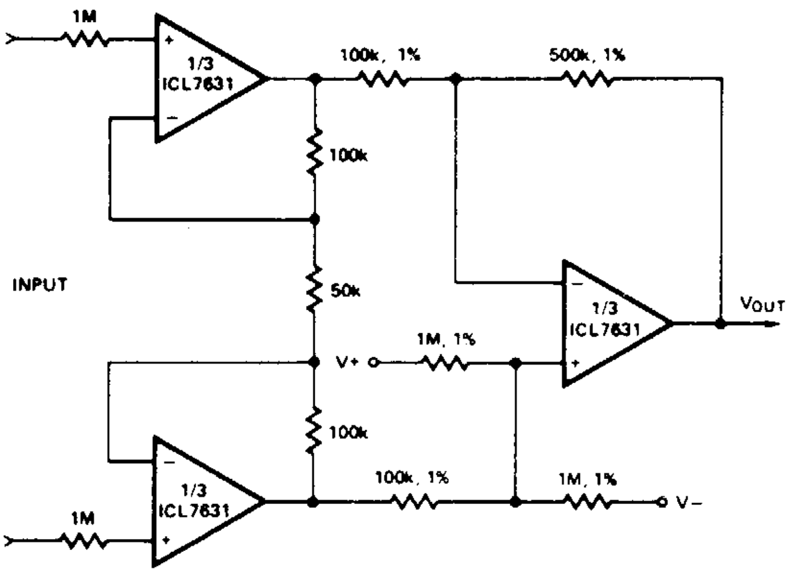
Fig. 55-1

HEART RATE MONITOR, Continued.

Circuit Notes

Light filtering through the finger tip is detected by the cadmium sulphide photoresistor CD1 which forms the feedback network for transducer amplifier section ICA producing a weak signal which is further amplified by ICB. This signal is now compared against a user adjusted threshold, comparator ICD triggers gating on the piezoelectric buzzer PZ1. On each falling edge of the comparator's output signal one-shot multivibrator ICD produces a 2 μs pulse which is inverted by Q1 and averaged by the RC network consisting of M1, C6 and C7. The 10 K trimpot R20 in Q1's collector circuit sets the scale factor for M_1 where full scale is 150 beats per minute.

MEDICAL INSTRUMENT PREAMPLIFIER



INTERSIL

Fig. 55-2

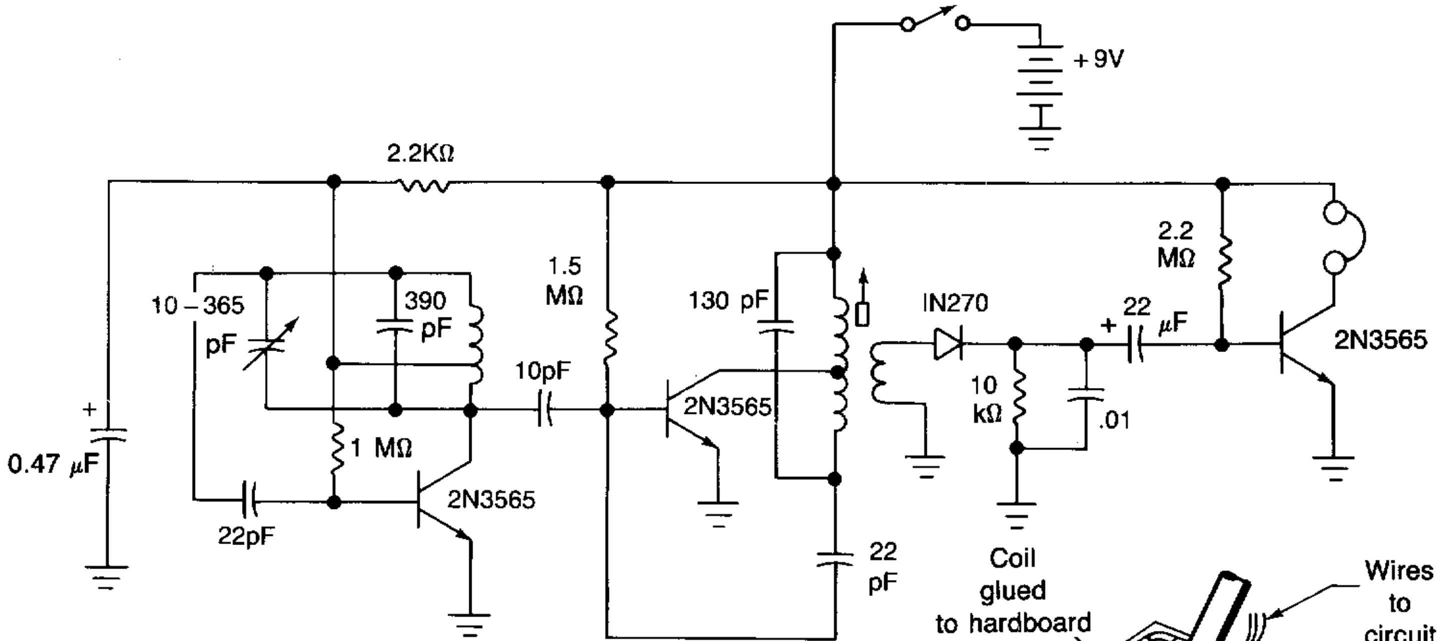
Circuit Notes

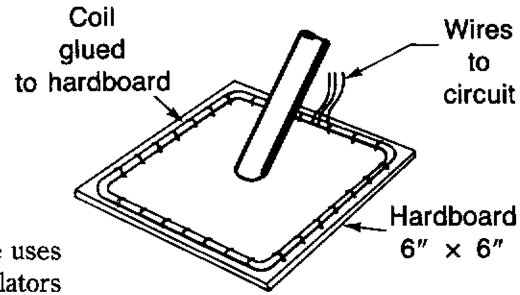
Note that A_{VOL} = 25; single Ni-cad battery operation. Input current (from sensors connected to the patient) is limited to less than 5 μA under fault conditions.

56
Metal Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

METAL LOCATOR II





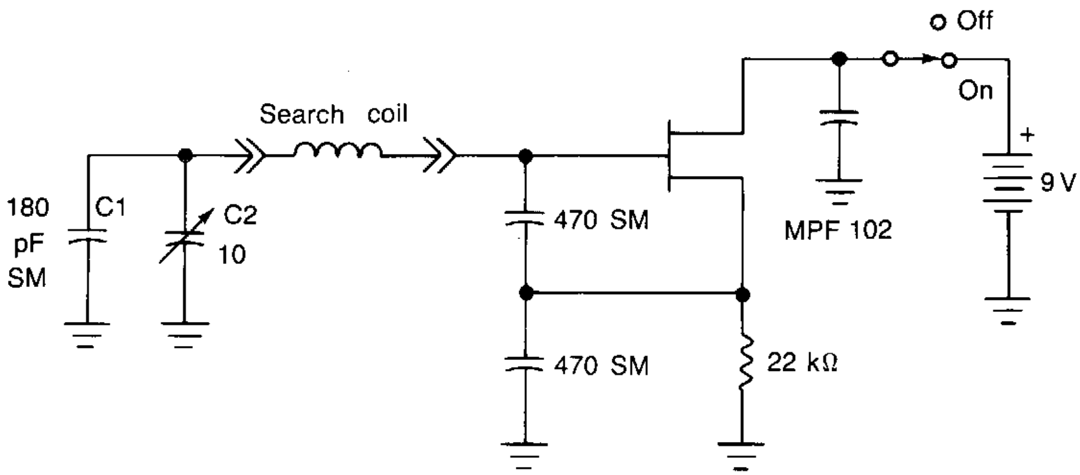
WILLIAM SHEETS

Fig. 56-1

Circuit Notes

The circuit consists of two oscillators, both working at about 465 kHz. One uses an if transformer and the other uses an inductor (the search coil L1). The oscillators are coupled by a capacitor (10 pF). A beat note (produced if the two oscillators are working closely together) is detected by the diode and fed to the headphone amplifier and the 22 μF capacitor. The search coil oscillator is tuned by a 10-365 pf variable capacitor. The search coil comprises 22 turns of wire (any gauge between 24 swg and 36 swg enamel) center tapped. The wire should be wound on a temporary form then taped and glued to a piece of hardboard. The coil size should be about 6″ × 6″. Headphones should be high impedance.

METAL LOCATOR



WILLIAM SHEETS **Fig. 56-2**

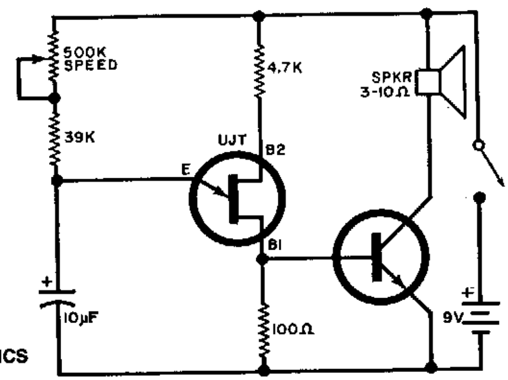
Circuit Notes

The search coil, C1 and C2 form a tuned circuit for the oscillator which is tuned near the center of the broadcast band. Tune a portable radio to a station near the middle of the band, then tune C2 until a squeal is heard as the two signals mix to produce a beat (heterodyne) note. Metal near the search coil will detune the circuit slightly, changing the pitch of the squeal. The search coil is 20 turns of number 30 enameled wire, wound on a 6″ × 8″ wood or plastic form. It is affixed at the end of a 30″ to 40″ wooden or plastic pole, and connected to the remainder of the metal detector circuit through a coaxial cable.

57
Metronomes

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SIMPLE METRONOME



POPULAR ELECTRONICS

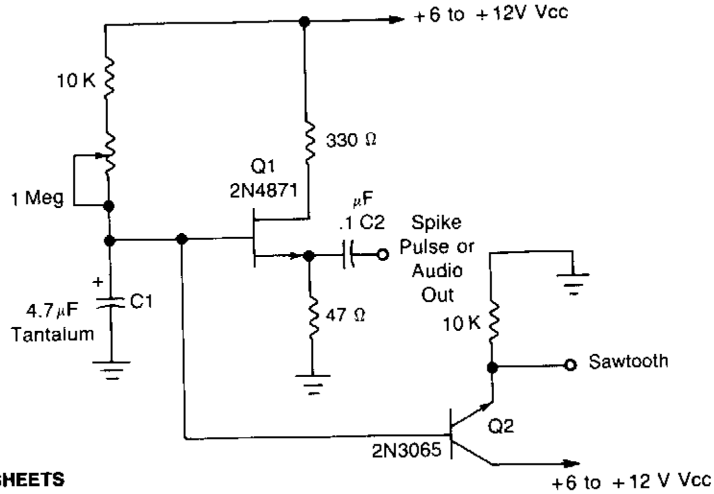
Fig. 57-1

Circuit Notes

Adjustable from 15 to 240 beats per minute. The UJT oscillator output is applied to a general purpose npn transistor which drives the speaker.

UJT = 2N4871
NPNxistor = TIP31

METRONOME I



WILLIAM SHEETS

Fig. 57-2

Circuit Notes

This simple oscillator uses a 2N4871 UJT to give pulses from 0.2 to about 20 Hz. A spike is available at C2, a sawtooth at the emitter of Q2 of about 2-3 V p-p, depending on V_{CC}.

AC-LINE OPERATED UNIJUNCTION METRONOME

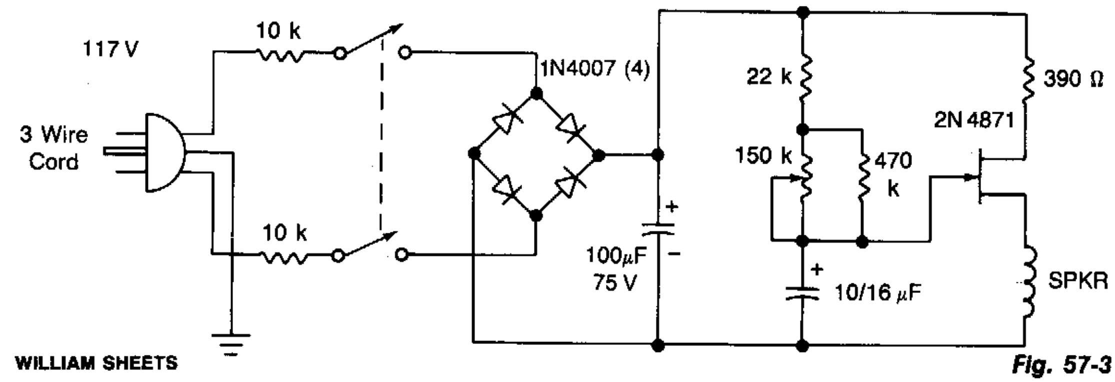


Fig. 57-3

Circuit Notes

The UJT-oscillator frequency is determined by the 100 μF capacitor and the effective resistance of the 22 K and 470 K resistors and the potentiometer. Rate can be varied from 42 to 208 beats/minute. The circuit should be housed in an insulated box for safety, or use ground (3-wire cord).

METRONOME II

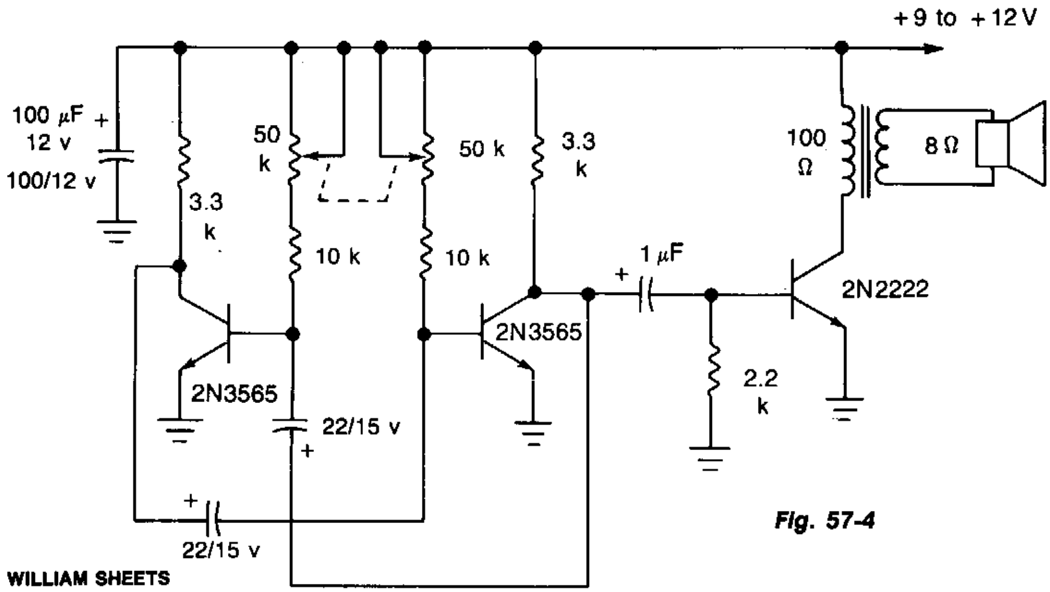


Fig. 57-4

Circuit Notes

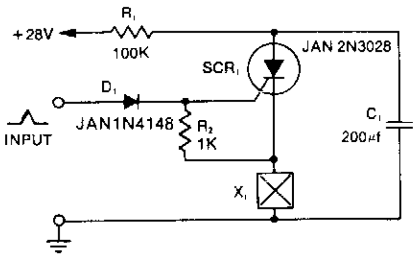
This simple circuit uses a multivibrator to generate the beats and a subsequent audio amplifier stage to increase the output level. Range of adjustment is approximately from 40 to 200 beats per minute set by the gauged potentiometer.

58 Miscellaneous Treasures

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Squib-Firing Circuit (I)
Squib-Firing Circuit (II)
Model Rocket Launcher
Push-On/Push Off Electronic Switch
Game Feeder Controller
Single LED Can Indicate Four Logic States
Inexpensive Radio-Control Uses Only SCR
Guitar and Bass Tuner
4-Channel Commutator
Two-Wire Tone Encoder
Differential Hold
5 MHz Phase-Encoded Data Read Circuitry
Shift Register
Power-On Reset
Noise Immune 60 Hz Line Sync
DC Static Switch (SCR Flip-Flop)

SQUIB-FIRING CIRCUIT (I)



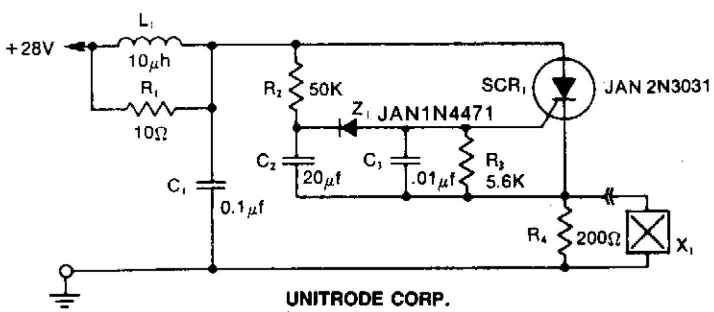
UNITRODE CORP.

Fig. 58-1

Circuit Notes

Capacitor C1 is charged to +28 V through R1 and stores energy for firing the squib. A positive pulse of 1 mA applied to the gate of SCR1 will cause it to conduct, discharging C1 into the squib load X1. With the load in the cathode circuit, the cathode rises immediately to +28 V as soon as the SCR is triggered on. Diode D1 decouples the gate from the gate trigger source, allowing the gate to rise in potential along with the cathode so that the negative gate-to-cathode voltage rating is not exceeded. This circuit will reset itself after test firing, since the available current through R1 is less than the holding current of the SCR. After C1 has been discharged, the SCR automatically turns off—allowing C1 to recharge.

SQUIB-FIRING CIRCUIT (II)



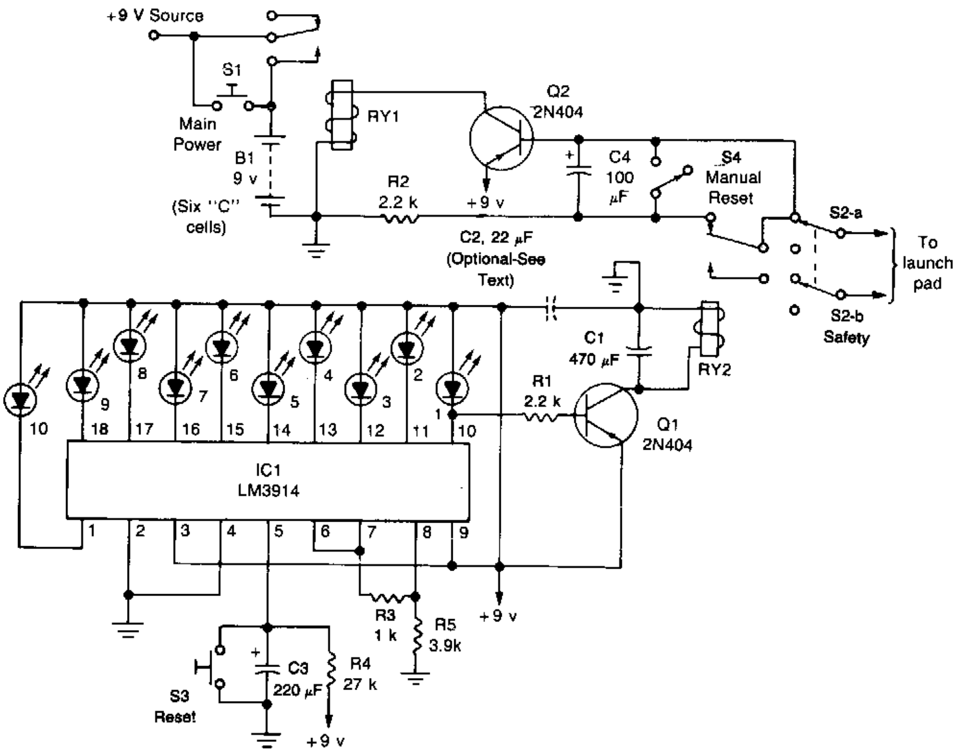
UNITRODE CORP.

Fig. 58-2

Circuit Notes

The LRC input network limits the anode dv/dt to a safe value—below 30 V/µs. R1 provides critical damping to prevent voltage overshoot. While a simple RC filter section could be used, the high current required by the squib would dictate a small value of resistance and a much larger capacitor. Resistor R3 provides dc bias stabilization, while C3 provides stiff gate bias during the transient interval when anode voltage is applied. The SCR is fired one second after arming by means of the simple R2C2Z1 time delay network. R4 provides a load for the SCR for testing the circuit with the squib disconnected—limiting the current to a level well within the continuous rating of the SCR. The circuit can be reset by opening the +28 V supply and then re-arming.

MODEL ROCKET LAUNCHER



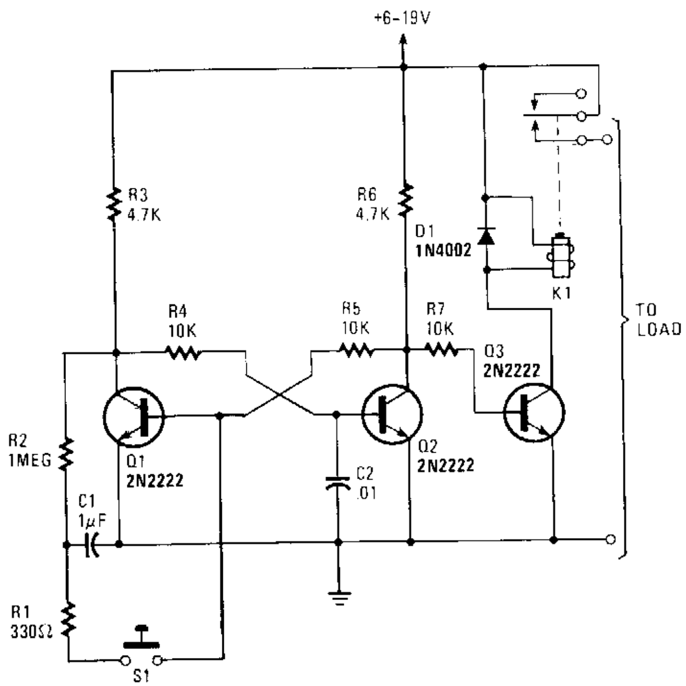
RADIO ELECTRONICS

Fig. 58-3

Circuit Notes

The circuit consists of the launch timer itself and an automatic-off timer. When power is applied to that IC, the countdown LED's sequence is on until they are all lit. When the last one LED1, is fully lit, transistor Q1 saturates, energizing RY2. When that happens, a circuit between the lantern battery at the launch pad and the nickel-chromium wire is completed; the wire heats up as before, and the rocket is launched. Resistor R4 and capacitor C3 determine the countdown timing; with the values shown it should be approximately 10 seconds. Resistors R3 and R5 set the LED brightness. Safety is of the utmost importance. That's the purpose of the second half of the circuit. When RY2 opens, the current flow to Q2 is disrupted. But, because of the presence of R2 and C4, the transistor remains saturated for about 3 seconds. After that, however, the transistor stops conducting and RY1 is de-energized. That cuts off the power to the rest of the circuit, and RY2 de-energizes again, breaking the circuit to the launch pad. Switch S3 is used to reset the countdown. Once that is done, pressing S1 starts the launch sequence; the rest is automatic. Switch S4 is used to latch RY1 manually if needed.

PUSH-ON/PUSH-OFF ELECTRONIC SWITCH



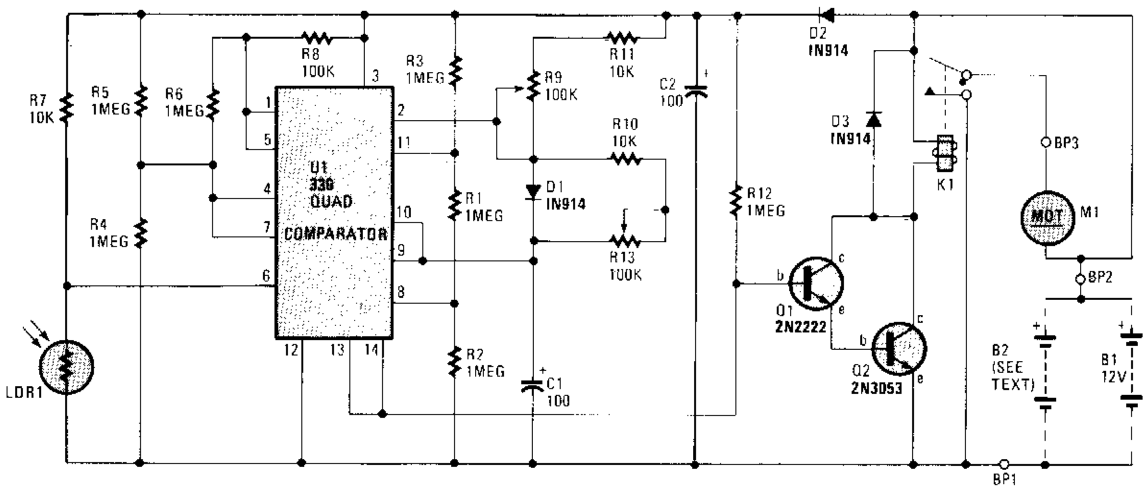
HANDS-ON ELECTRONICS

Fig. 58-4

Circuit Notes

Transistors Q1 and Q2 make up the flip-flop while Q3 drives a reed relay. When power is first applied to the circuit, Q1 and Q3 are conducting and Q2 is cut off. Momentarily closing S1 causes the flip-flop to switch states—Q1 cuts off and Q2 conducts. When Q2 is conducting, its collector drops to around 0.6 volt. That prevents base current from flowing into Q3 so it is cut off, de-energizing relay K1. The flip-flop changes state every time S1 is pressed. Capacitors C1 and C2 ensure that Q1 is always the transistor that turns on when power is first applied to the circuit. When power is first applied to the basic flip-flop, the initial status is random—Q1 and Q2 both try to conduct and, usually, the transistor with the higher gain will take control, reaching full conduction and cutting off the other one. However, differences in the values of the collector and coupling resistors will also influence the initial state at power-on. With C2 in the circuit, it and R4 form an R-C network that slightly delays the rise in Q2's base voltage. That gives Q1 sufficient time to reach saturation and thus take control.

GAME FEEDER CONTROLLER

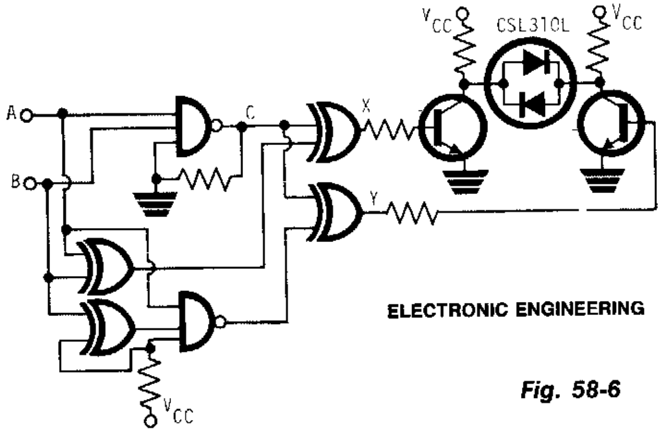


HANDS-ON ELECTRONICS *Fig. 58-5*

Circuit Notes

The circuit is built around an LM339 quad comparator, U1, which forms the basis of a Schmitt trigger, timer circuit, and a window comparator. One comparator within the LM339 (pins 1, 7, 6), plus LDR1, R4, R5, R6 and R8, is used as a Schmitt trigger. The timer circuit (which receives its input from the Schmitt trigger) consists of R9, R10, R11, R13. The last two-fourth's of U1 (pins 8, 9, 10, 11, 13 and 14) are wired as a window comparator. The two inputs to the window comparator are derived from the charge on capacitor C1—which is fed to pins 9 and 10 of U1. The other inputs are picked from two points along a voltage-divider network, consisting of R1, R2, and R3. Diode D1 is used as a blocking diode, forcing capacitor C1 to discharge through R10 and R13. The window comparator looks for any voltage falling between one-third and two-thirds of the supply voltage. When the voltage falls between those two points, the output of the window comparator (pins 13/14) goes high. Transistors Q1, and Q2 are turned on, when the pins 13/14 junction goes high, energizing the relay, K1. The energized relay provides a dc path to ground, activating the motor, M1, which reloads the feeder. The timer circuit also provides immunity from triggering, due to lightning. The on-time of relay K1 is determined by the charge cycle of C1, R11, and R9 or the discharge cycle of C1, R10, and R13. Changing the value of either a resistor or the capacitor, changes the timing cycle.

SINGLE LED CAN INDICATE FOUR LOGIC STATES



ELECTRONIC ENGINEERING

Fig. 58-6

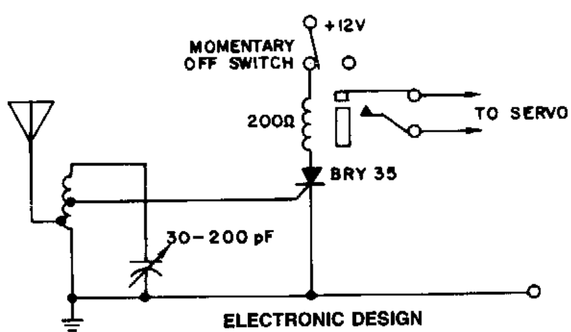
Circuit Notes

The LED is the CSL310L which contains a red LED and a green LED connected back to back and mounted close together in a single moulding. The LED can emit red or green light by controlling the polarity of the applied voltage and if the polarity is switched at a rate of several hundred Hertz the emitted light appears yellow. The four combinations of inputs A and B can therefore be converted to four LED states—red, green, yellow and off. The truth table shows the LED colors corresponding to the combinations of A and B levels.

Truth Table

A	B	X	Y	LED color
0	0	1	0	red
0	1	0	0	off
1	0	0	1	green
1	1	C	C	yellow

INEXPENSIVE RADIO-CONTROL USES ONLY ONE SCR



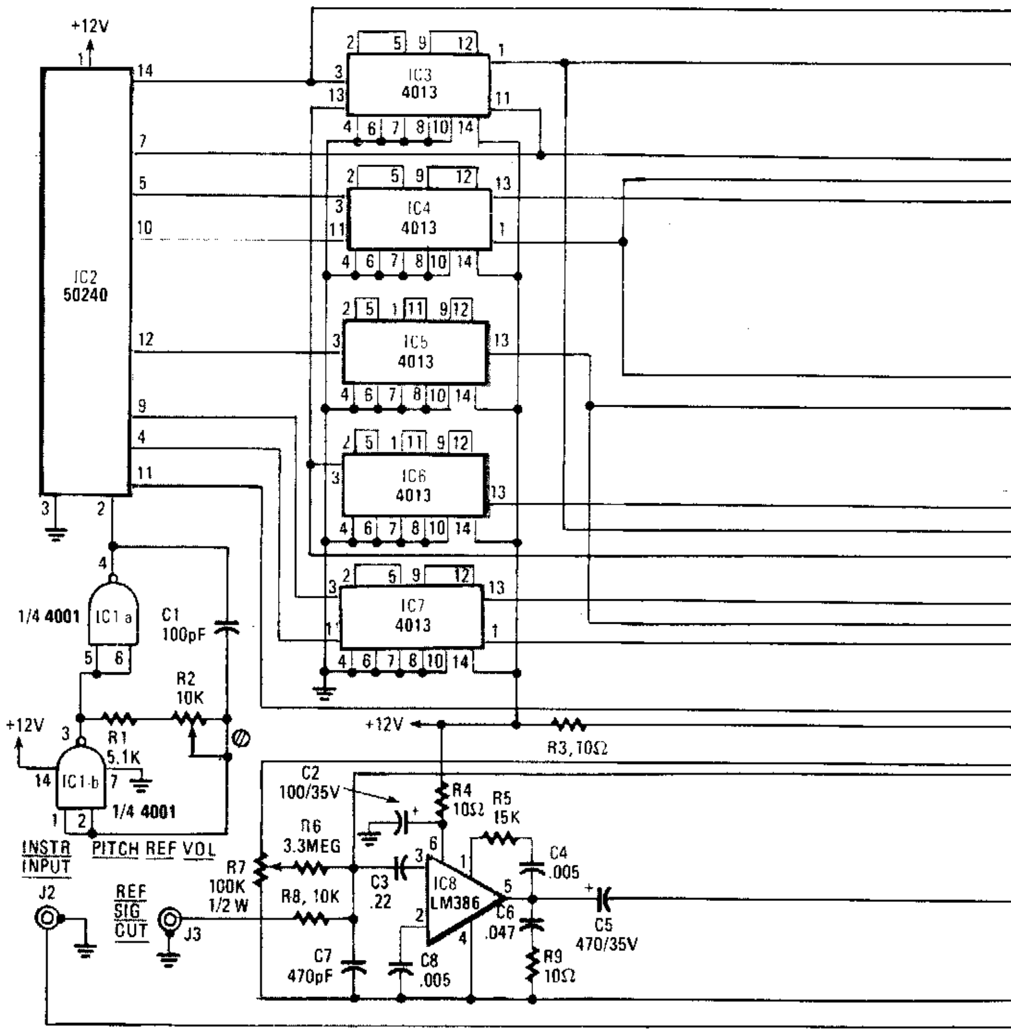
ELECTRONIC DESIGN

Fig. 58-7

Circuit Notes

A simple and effective receiver for actuating garage doors, alarms, warning systems, etc. The SCR, which has a very low trigger current 30 μA is typical—it requires an input power of only 30 μW to activate the relay. A high Q tuned antenna circuit assures rejection of spurious signals. A whip or wire antenna is adequate up to 100 feet from a low power transistor transmitter. A momentary-off switch resets the circuit.

GUITAR AND BASS TUNER

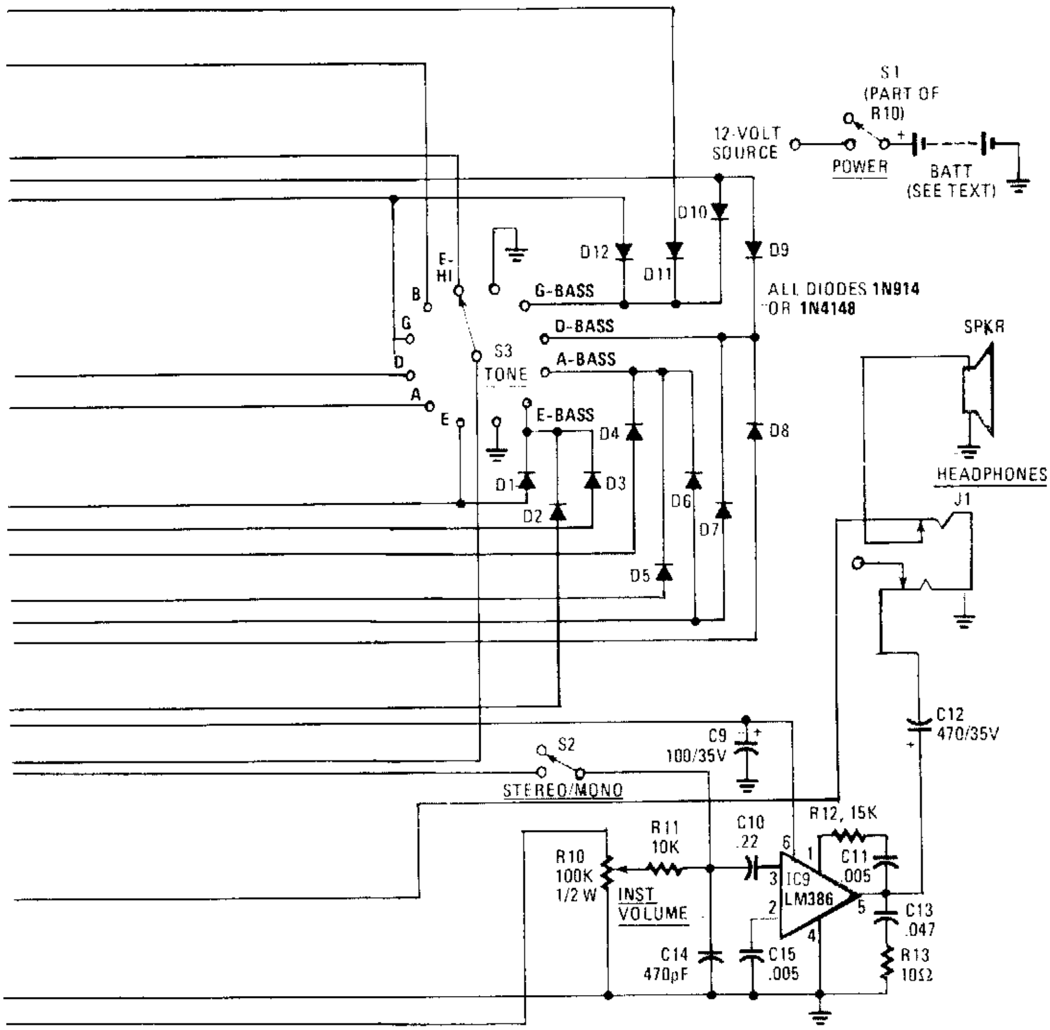


RADIO-ELECTRONICS

Fig. 58-8

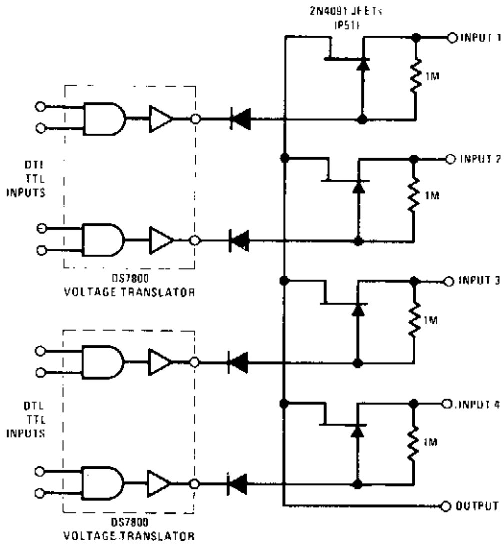
Circuit Notes

The heart of the circuit is IC2, a 50240 top-octave generator. That device uses a single input-frequency to generate all twelve notes of the musical scale. The input signal is provided by IC1, a 4001 quad 2-input NOR gate. Two sections of that IC are used to form an oscillator that runs at approximately 2 MHz. The frequency can be adjusted by trimmer potentiometer R2. Dual D flip-flops, IC3-IC7, are used as frequency dividers. They divide down the upper-octave frequencies from IC2, thus generating the lower-frequency notes required for the pitch references. The chords for the bass pitch-references are composed of three notes each. Those notes are taken from various outputs



of IC2-IC7 through isolation diodes D1-D12. All signals are routed to the TONE switch, S3. The wiper arm of that switch is connected through R7 to the input of audio power-amplifier IC8, an LM386. The resistor acts as a volume control for the pitch reference. Another LM386, IC9, serves as an amplifier for the instrument being tuned, with R10 acting as its volume control. The outputs of IC8 and IC9 are coupled, through C5 and C12 respectively, to the headphone jack, J1. Switch S2 STEREO/MONO is used to mix the reference and instrument signals at IC9 for mono operation. Power is supplied by eight "AA" cells connected in series.

4-CHANNEL COMMUTATOR



Circuit Notes

This 4-channel commutator used the 2N4091 to achieve low channel on resistance (<30 ohm) and low off current leakage. The DS7800 voltage translator is a monolithic device that provides from 10 V to −20 V gate drive to the JFETs while at the same time providing DTL/TTL logic compatibility.

Fig. 58-9

NATIONAL SEMICONDUCTOR CORP.

TWO-WIRE TONE ENCODER

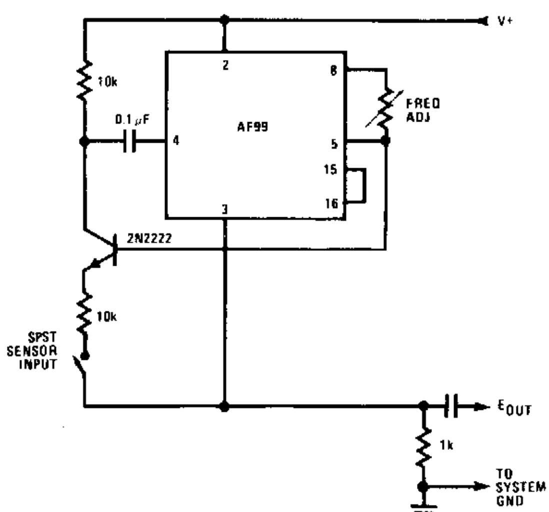


Fig. 58-10

NATIONAL SEMICONDUCTOR CORP.

DIFFERENTIAL HOLD

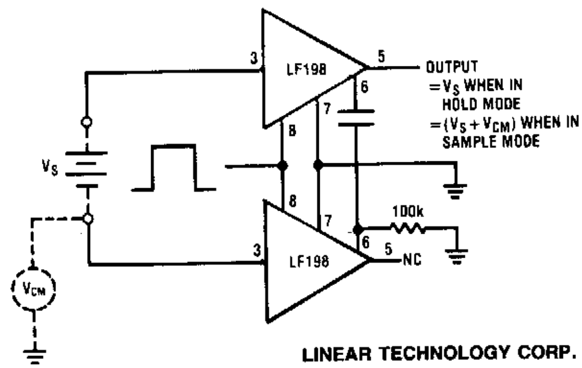
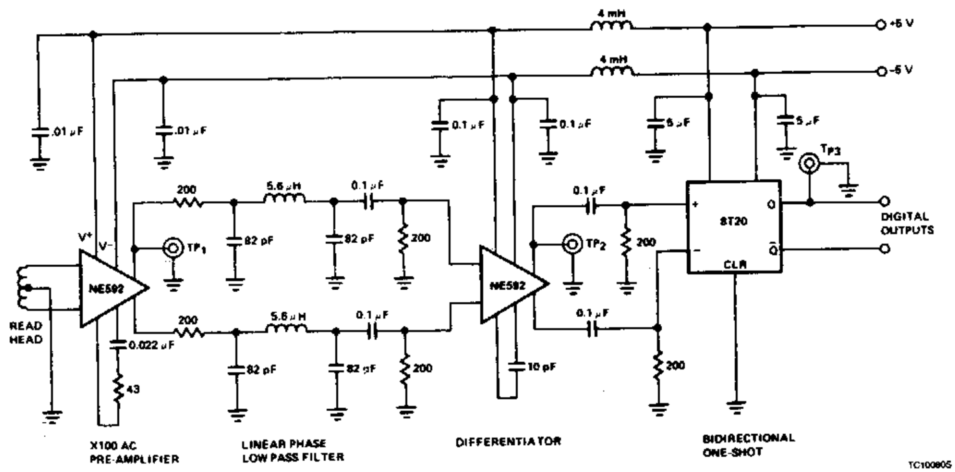


Fig. 58-11

LINEAR TECHNOLOGY CORP.

5 MHz PHASE—ENCODED DATA READ CIRCUITRY



NOTE:
All resistor values are in ohms

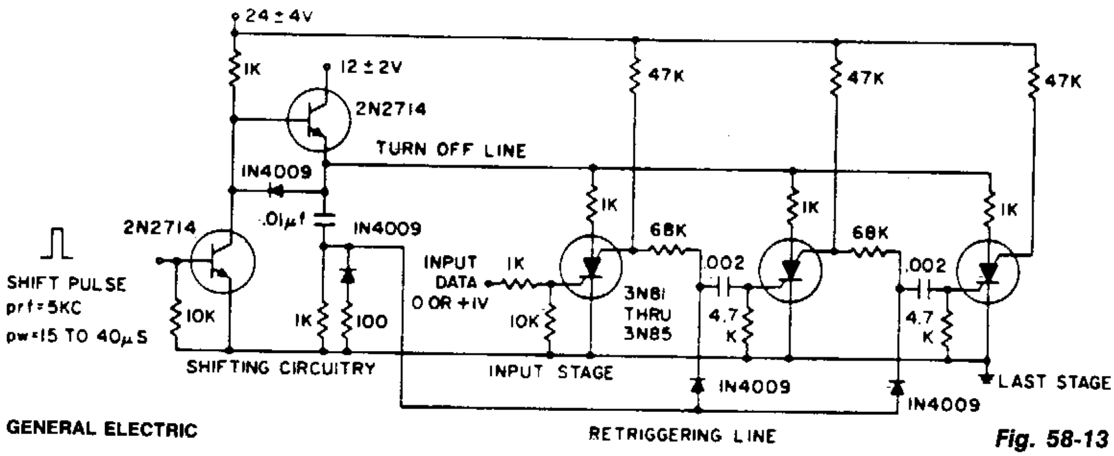
SIGNETICS

Fig. 58-12

Circuit Notes

Readback data is applied directly to the input of the first NE592. This amplifier functions as a wide-band ac coupled amplifier with a gain of 100. By direct coupling of the readback head to the amplifier, no matched terminating resistors are required and the excellent common-mode rejection ratio of the amplifier is preserved. The dc components are also rejected because the NE592 has no gain at dc due to the capacitance across the gain select terminals. The output of the first stage amplifier is routed to a linear phase shift low-pass filter, with a characteristic impedance of 200 ohms. The second NE592 is utilized as a low noise differentiator/amplifier stage. The output of the differentiator/amplifier is connected to the 8T20 bidirectional monostable unit to provide the proper pulses at the zero-crossing points of the differentiator.

SHIFT REGISTER



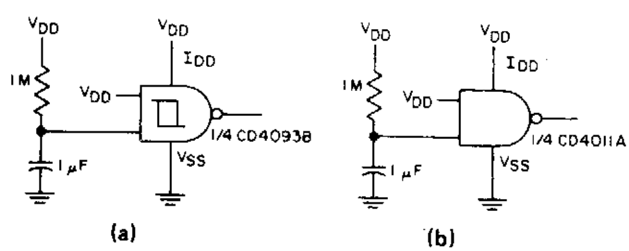
GENERAL ELECTRIC

Fig. 58-13

Circuit Notes

The shift pulse amplitude is less than 15 volts. If a stage is off, the shift pulse will not be coupled to the next stage. If it is on, the diode will conduct triggering the next stage. Just prior to the shift pulse the anode supply is interrupted to turn off all stages. The stored capacitor charge determines which stages will be retriggered.

POWER-ON RESET



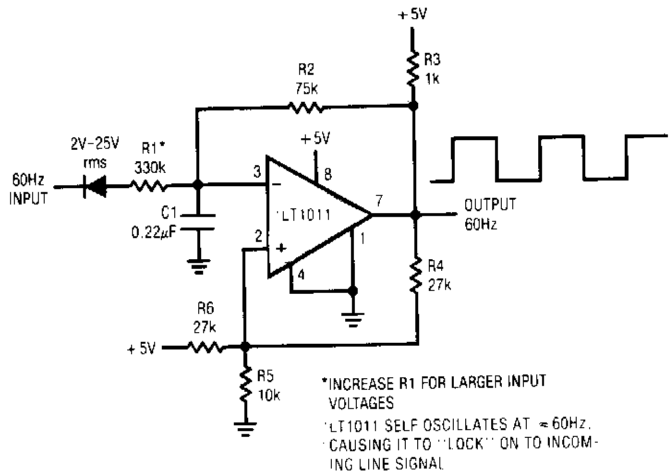
RCA

Fig. 58-14

Circuit Notes

A reset pulse is often required at power-on in a digital system. This type of reset pulse is ideally provided by this circuit. Because of the high input impedance of the Schmitt trigger, long reset pulse times may be achieved without the excess dissipation that results when both output devices are on simultaneously, as in an ordinary gate device (B).

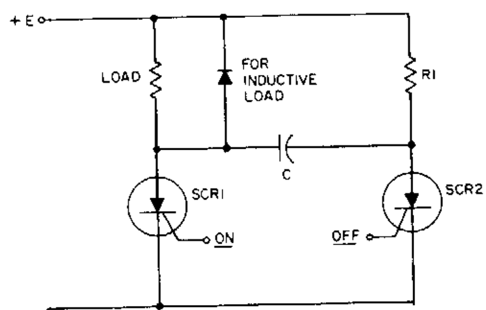
NOISE IMMUNE 60 Hz LINE SYNC



LINEAR TECHNOLOGY CORP.

Fig. 58-15

DC STATIC SWITCH (SCR FIIP-FLOP)



GENERAL ELECTRIC

Fig. 58-16

Circuit Notes

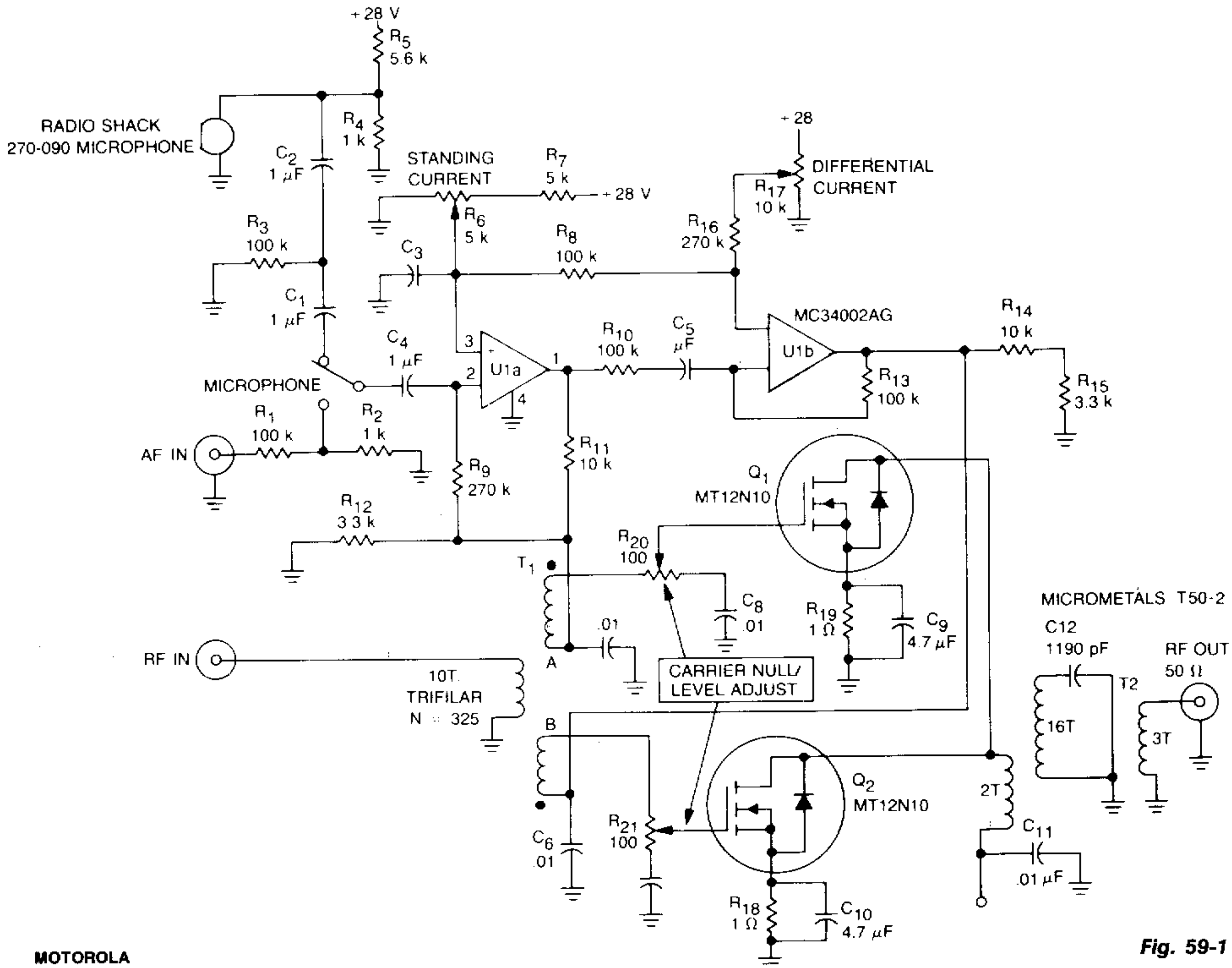
This circuit is a static SCR switch for use in a dc circuit. When a low power signal is applied to the gate of SCR1, this SCR is triggered and voltage is applied to the load. The right hand plate of C charges positively with respect to the left hand plate through R1. When SCR2 is triggered on, capacitor C is connected across SCR1, so that this SCR is momentarily reverse biased between anode and cathode. This reverse voltage turns SCR1 off provided the gate signal is not applied simultaneously to both gates. The current through the load will decrease to zero in an exponential fashion as C becomes charged.

59

Modulator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DOUBLE SIDEBAND, SUPPRESSED CARRIER RF MODULATOR



MOTOROLA

Fig. 59-1

DOUBLE SIDEBAND, SUPPRESSED CARRIER RF MODULATOR , Continued.

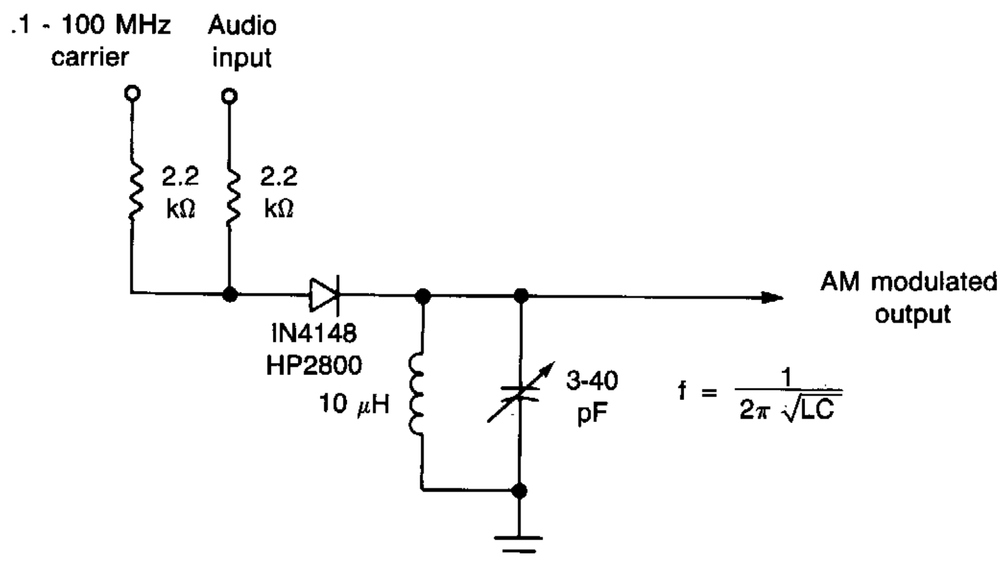
Circuit Notes

An RF input is applied to the primary of T1, which applies equal amplitude, opposite phase RF drive for output FETs Q1 and Q2. With no AF modulation at points A and B, the opposite phase RF signals cancel each other and no output appears at the 50 V output connector.

When AF modulation is applied to points A and B, a modulated RF output is obtained. The dc stability and low frequency gain are improved by source resistors R18 and R19.

A phase inverter consisting of a dual op amp (U1a and U1b) produces the out-of-phase, equal amplitude AF modulation signals.

LOW-DISTORTION LOW-LEVEL AMPLITUDE MODULATOR



$$f = \frac{1}{2\pi\sqrt{LC}}$$

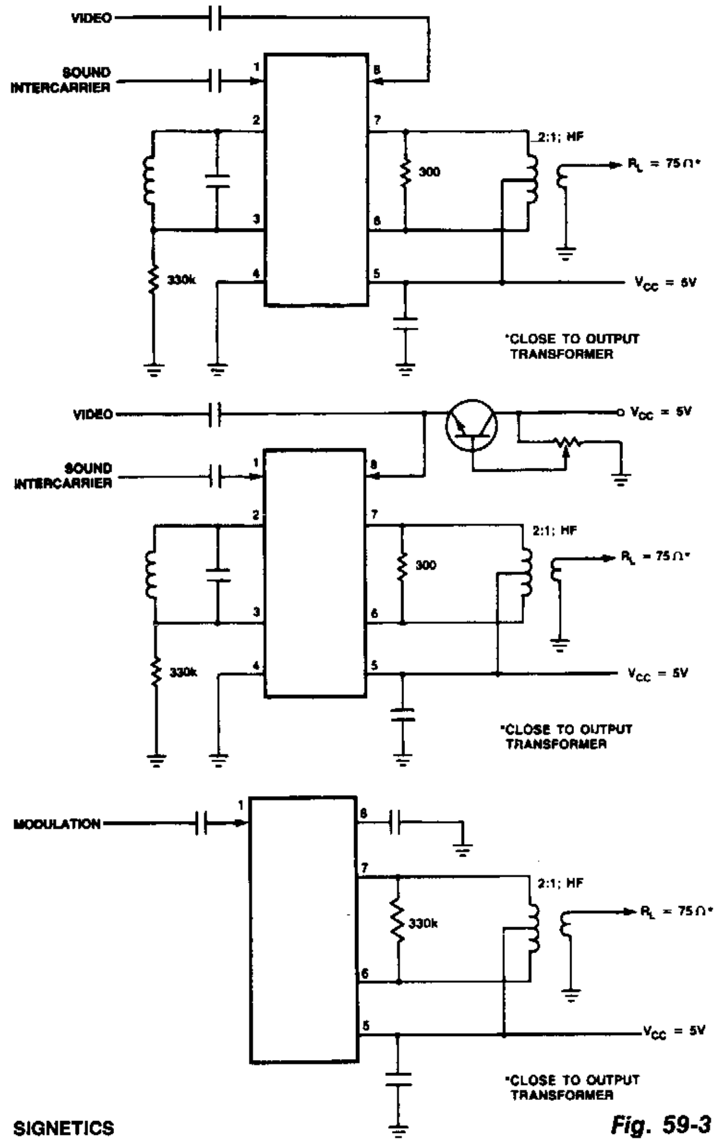
WILLIAM SHEETS

Fig. 59-2

Circuit Notes

This simple diode modulator delivers excellent results when used for high percentage modulation at low signal levels. Constants are shown for a carrier frequency of about 10 MHz, but, with a suitable tank, the circuit will give good results at any frequency at which the diode approximates a good switch. To extend frequency above that for which the IN4148 is suited, a hot-carrier diode (HP2800, etc.) can be substituted. A shunt resistor across the tank circuit can be used to reduce the circuit Q so as to permit high percentage modulation without appreciable distortion.

VIDEO MODULATOR CIRCUIT

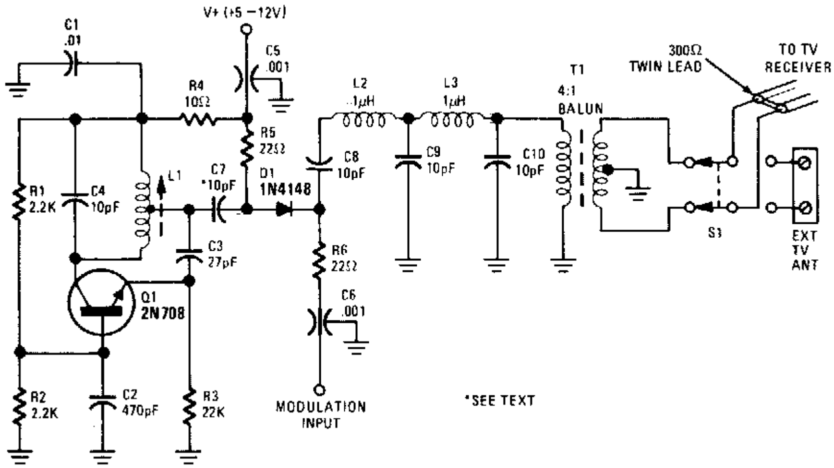


SIGNETICS *Fig. 59-3*

Circuit Notes

These are modulator circuits for modulation of video signals on a VHF/UHF carrier. The circuits require a 5 V power supply and few external components for the negative modulation mode. For positive modulation an external clamp circuit is required. The circuits can be used as general-purpose modulators without additional external components. The IC is TDA6800.

VIDEO MODULATOR

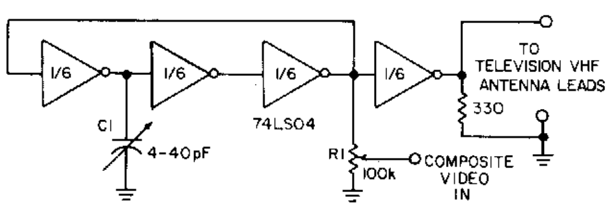


RADIO-ELECTRONICS *Fig. 59-4*

Circuit Notes

This circuit permits direct connection of composite video signals from video games and microcomputers to the antenna terminals of TV sets. The output signal level is controlled by the modulation input.

TTL OSCILLATOR INTERFACES DATA FOR DISPLAY BY A TELEVISION SET



ELECTRONIC DESIGN *Fig. 59-5*

Circuit Notes

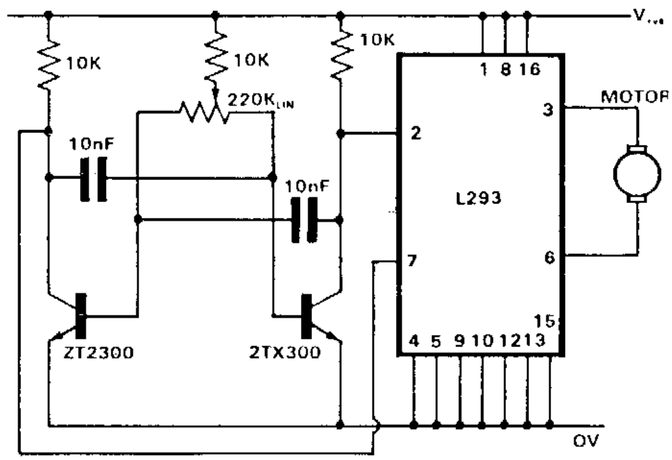
Three gates of a 74LS04 form the oscillator circuit. Capacitor C1 allows fine-frequency adjustment to a specific television channel and helps stabilize the circuit. Potentiometer R1 acts as the mixing input and provides adjustment of the contrast ratio for the best viewing. A fourth gate buffers and helps stabilize the oscillator.

60

Motor Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

BI-DIRECTIONAL PROPORTIONAL MOTOR CONTROL



ELECTRONIC ENGINEERING *Fig. 60-1*

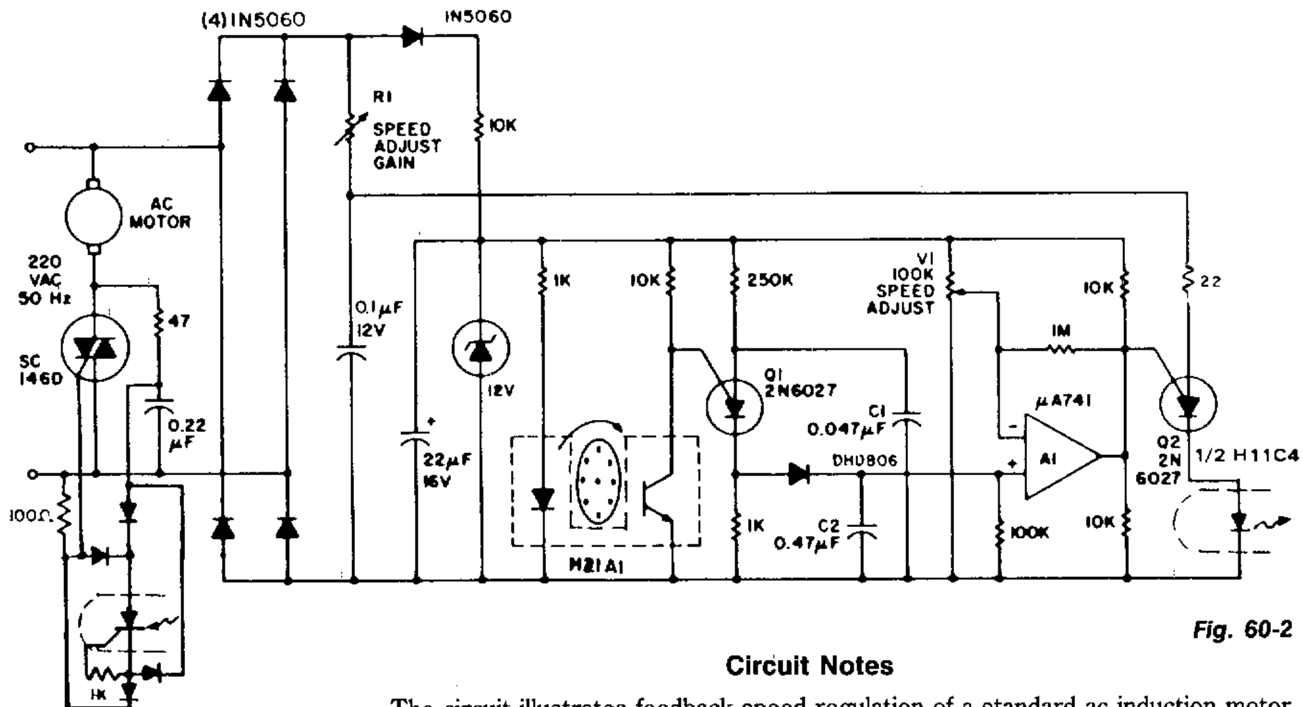
Circuit Notes

The control of both direction and of proportional motor speed is achieved by rotation of a single potentiometer. The motor driver is an SGS integrated circuit L293 which will drive up to 1 amp in either direction, depending on the logic state of input 1 and input 2 as per table.

I/P 1	I/P 2	Function
High	Low	Motor turns one way
Low	High	Motor reverses

By applying a variable M/S ratio flip-flop to these inputs, both speed and direction will be controlled. With RV1 in its center position the M/S will be 1:1 whereby the motor will remain stationary due to its inability to track at the flip-flop frequency. Movement of RV1 in either direction will gradually alter the M/S ratio and provide an average voltage bias in one direction proportional to the M/S ratio.

AC MOTOR CONTROL



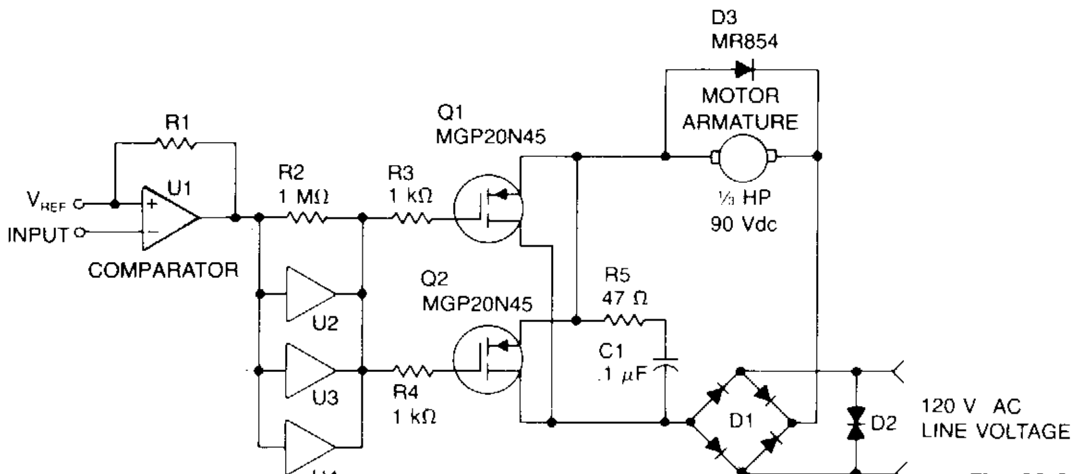
GENERAL ELECTRIC

Fig. 60-2

Circuit Notes

The circuit illustrates feedback speed regulation of a standard ac induction motor, a function difficult to accomplish other than with a costly, generator type, precision tachometer. When the apertured disc attached to the motor shaft allows the light beam to cross the interrupter module, the programmable unijunction transistor, Q1, discharges capacitor, C1, into the much larger storage capacitor, C2. The voltage on C2 is a direct function of the rotational speed of the motor. Subsequently, this speed-related potential is compared against an adjustable reference voltage, V1, through the monolithic operational amplifier, A1, whose output, in turn, establishes a dc control input to the second P.U.T. (Q2). This latter device is synchronized to the ac supply frequency and furnishes trigger pulses in the conventional manner to the triac at a phase angle determined by the speed control, R1, and by the actual speed of the motor.

PWM MOTOR SPEED CONTROL



MOTOROLA

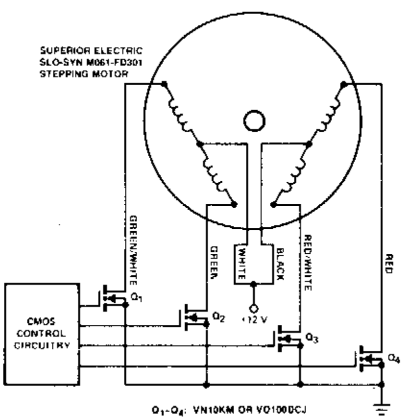
Fig. 60-3

Circuit Notes

Speed control is accomplished by pulse width modulating the gates of two MGP20N45 TMOS devices. Therefore, motor speed is proportional to the pulse width of the incoming digital signal, which can be generated by a microprocessor or digital logic.

The incoming signal is applied to comparator U1, then to paralleled inverters U2, U3, and U4 that drive the two TMOS devices, which, in turn, control power applied to the motor armature. Bridge rectifier D1 supplies fullwave power that is filtered by R5 and C1. Free-wheeling diode D3 (MR854) prevents high voltage across Q1 and Q2. A back-to-back zener diode, D2, protects against transients and high voltage surges.

STEPPING MOTOR DRIVER



SILCONIX, INC.

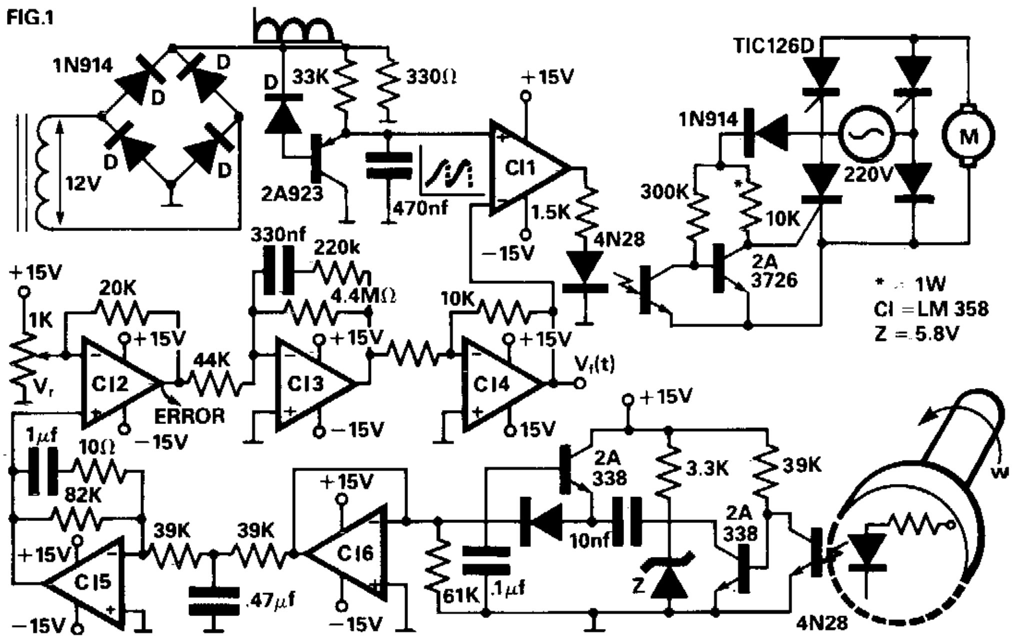
Fig. 60-4

Circuit Notes

Stepping motors find wide use in disk drives and machine control. MOSPOWER transistors are ideal motor drivers because of their freedom from second breakdown. Note that snubbing networks are not used because load line shaping is not necessary with MOSPOWER and the inductance of the motor is fairly low so that the inductive spike is small. The MOSFET gates are tied directly to the outputs of the CMOS control circuitry. The logic is arranged to sequence the motor in accordance with the needs of the application.

LOW-COST SPEED REGULATOR FOR DC MOTORS

FIG.1



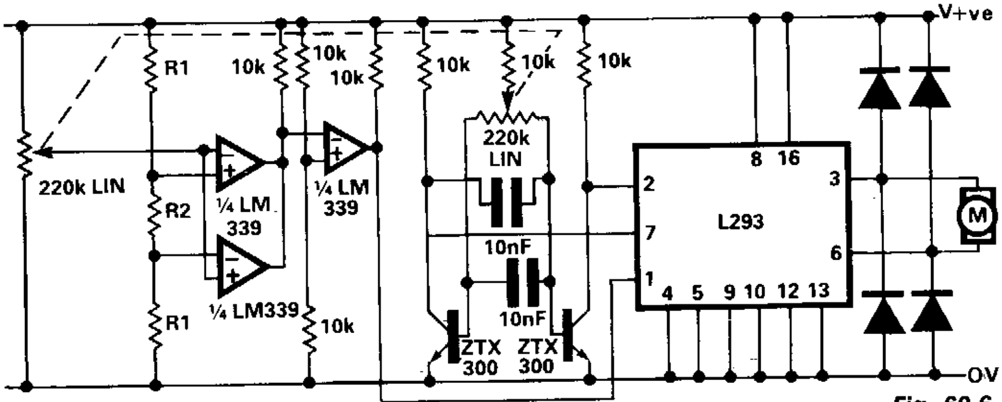
ELECTRONIC ENGINEERING

Fig. 60-5

Circuit Notes

A four thyristor controlled bridge is used for operation in two quadrants of the torque-speed characteristics. In the trigger circuits the usual pulse transformers were replaced by self biased circuits which minimize gate power consumption and increase noise immunity. Electrical isolation is guaranteed by the use of optocouplers. The trigger pulses are generated by the comparison between an error signal, previously processed and amplified, and a line synchronism signal. The converter's output is a dc voltage proportional to the speed, which after being compared with a reference signal, becomes the error signal.

MOTOR SPEED CONTROL CIRCUIT



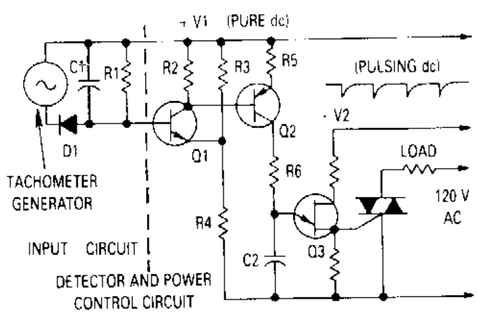
ELECTRONIC ENGINEERING

Fig. 60-6

Circuit Notes

A shortcoming of the above bi-directional proportional motor control circuit is that with the potentiometer in its center position the motor does not stop, but creeps due to the difficulty in setting the potentiometer for an exact 1:1 mark-space ratio from the flip-flop. This modified circuit uses a second potentiometer, ganged with the first used to inhibit drive to the motor near the center position. This potentiometer is connected between the supply lines and feeds a window comparator which in turn drives the inhibit input of the L293.

CONSTANT SPEED MOTOR CONTROL USING TACHOMETER FEEDBACK



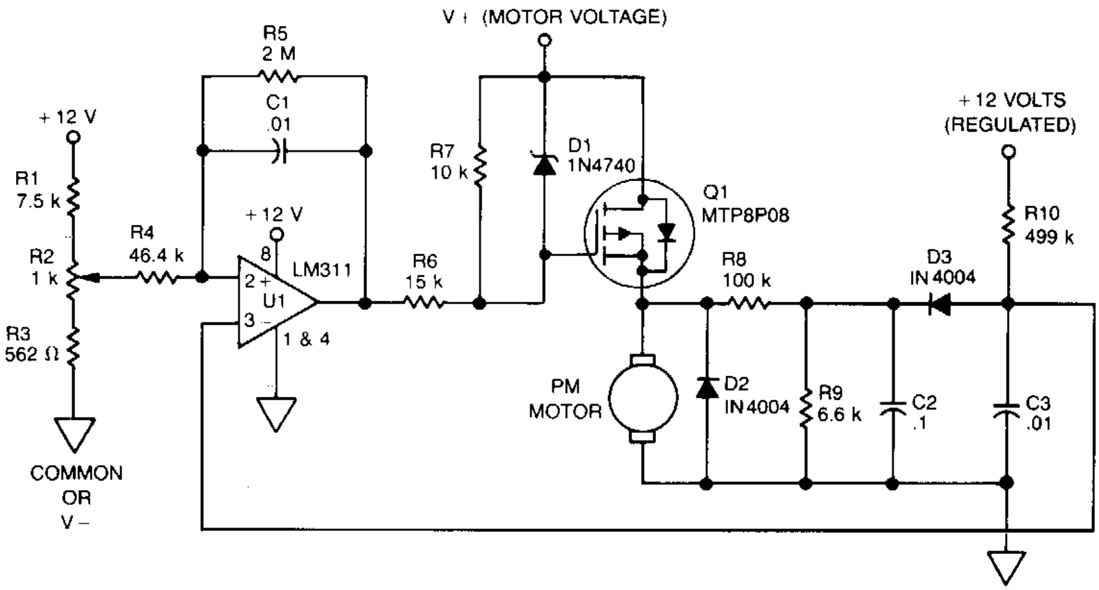
MOTOROLA

Fig. 60-7

Circuit Notes

The generator output is rectified then filtered and applied between the positive supply voltage and the base of the detector transistor. This provides a negative voltage which reduces the base-voltage when the speed increases. In normal operation, if the tachometer voltage is less than desired, the detector transistor is turned on, then turns on Q2 which causes the timing capacitor for the unijunction transistor to charge quickly. As the tachometer output approaches the voltage desired, the base-emitter voltage is reduced to the point at which Q1 is almost cut off. Thereby, the collector current which charges the unijunction timing capacitor is reduced, causing that capacitor to charge slowly and trigger the thyristor later in the half cycle. In this manner, the average power to the motor is reduced until just enough power to maintain the desired motor speed is allowed to flow.

BACK EMF PM MOTOR SPEED CONTROL



MOTOROLA

Fig. 60-8

Circuit Notes

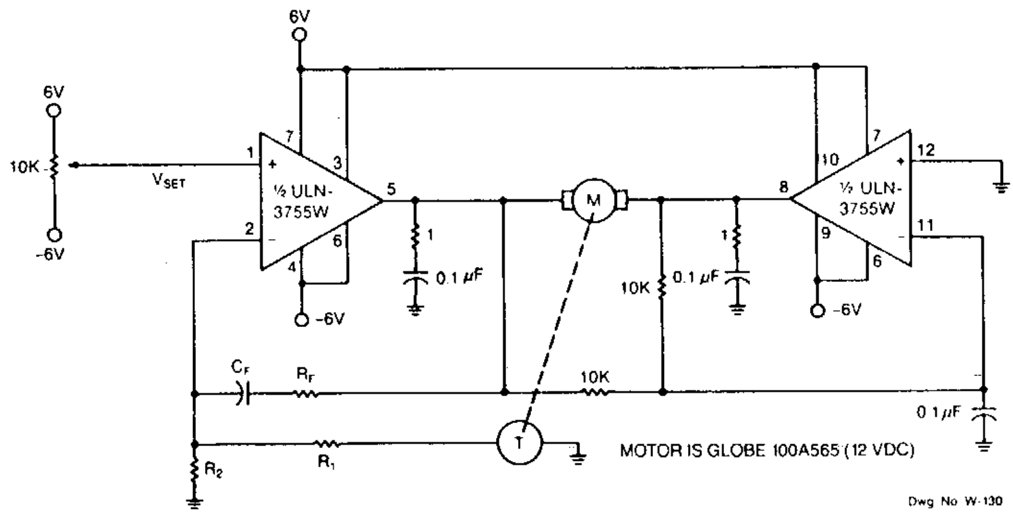
The use of power MOSFETs allows a direct interface between logic and motor power, which permits circuit simplicity as well as high efficiency. This speed control circuit can be packaged on a 22-pin, double-sided, 3.5 × 4-in. pc board.

A 12 V control supply and a TRW BL11, 30 V motor are used; with minor changes other motor and control voltages can be accommodated. For example, a single 24 V rail could supply both control and motor voltages. Motor and control voltages are kept separate here because CMOS logic is used to start, stop, reverse and oscillate the motor with a variable delay between motor reversals.

Motor speed is established by potentiometer R2, which applies a corresponding dc voltage to the + input of comparator U1, whose output is then applied to TMOS device MTP8P08 (Q1). Zener diode D1 limits the drive to Q1. The output of Q1 drives the permanent magnet motor.

Back emf is obtained from the motor via the network consisting of R8, R9, R10, C2, C3 and D3; it is applied to—input of comparator U1.

DC MOTOR SPEED CONTROL



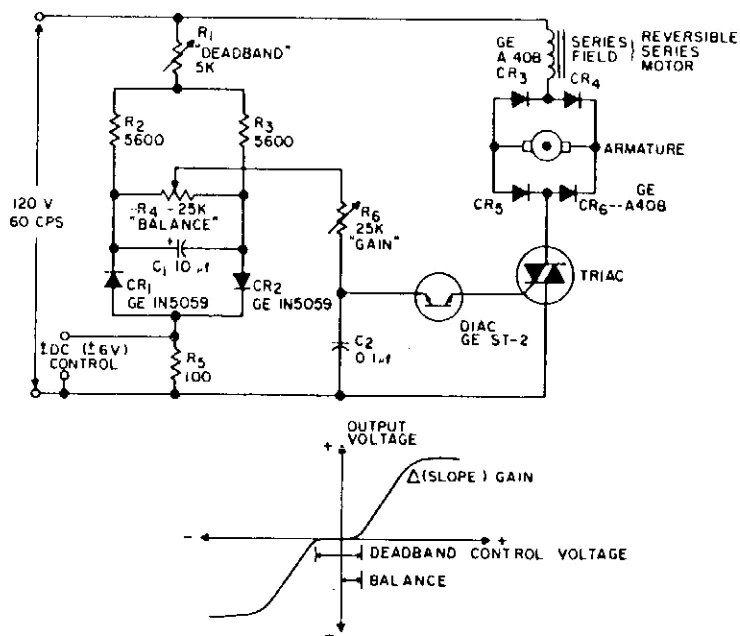
SPRAGUE ELECTRIC CO.

Fig. 60-9

Circuit Notes

Power op amps provide accurate speed control for dc motors. The circuit provides bidirectional speed control. The amplifiers' push-pull configuration ensures a full rail-to-rail voltage swing (minus the output stages' saturation drops) across the motor in either direction. The circuit uses a mechanically-coupled tachometer to provide speed-stabilizing feedback to the first amplifier section. The motor's speed and direction of rotation is set by adjusting the 10 k ohm potentiometer at the amplifier's noninverting input. The RFCF feedback network prevents oscillation by compensating for the inherent dynamic mechanical lag of the motor. Select the RFCF time constant to match the particular motor's characteristics.

REVERSING MOTOR DRIVE, DC CONTROL SIGNAL



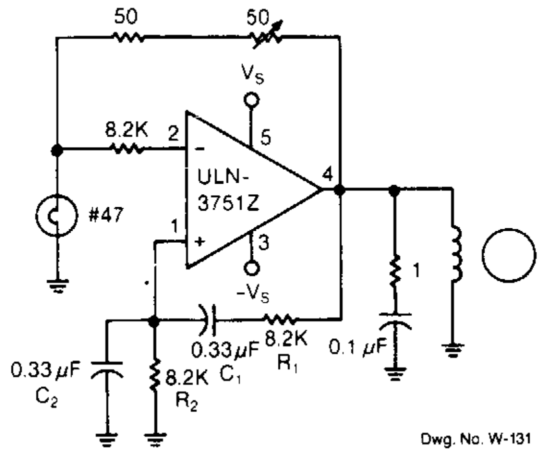
GENERAL ELECTRIC *Fig. 60-10*

Circuit Notes

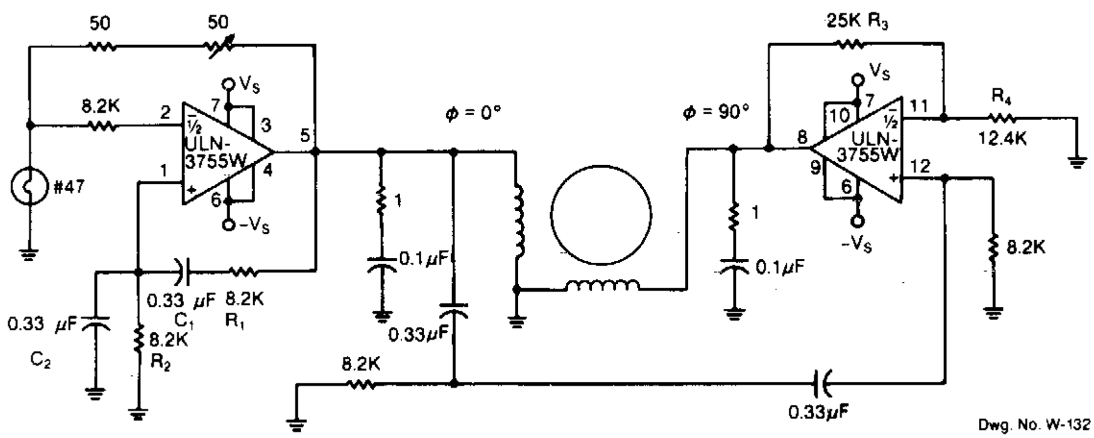
This is a positioning servo drive featuring adjustment of balance, gain, and deadband. In addition to control from a dc signal, mechanical input can be fed into the balance control, or that control could be replaced by a pair of resistance transducers for control by light or by temperature.

N-PHASE MOTOR DRIVERS

SINGLE-PHASE AC MOTOR DRIVER



TWO-PHASE AC MOTOR DRIVER



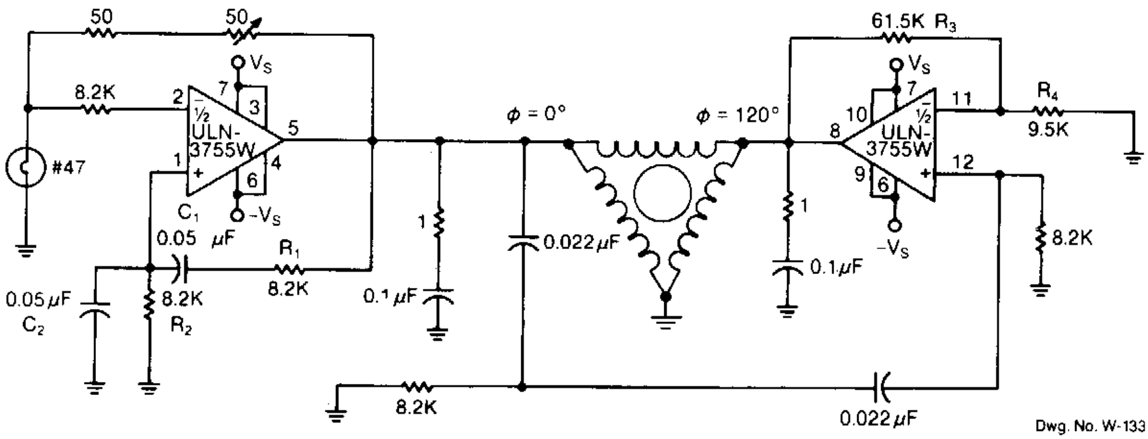
SPRAGUE ELECTRIC CO.

Fig. 60-11

Circuit Notes

Because of its high amplification factor and built-in power-output stage, an integrated power operational amplifier makes a convenient driver for ac motors. One op amp can be configured as an oscillator to generate the required ac signal. The power-output stage, of course, supplies the high-current drive to the motor. The controlling op amp is

THREE-PHASE AC MOTOR DRIVER

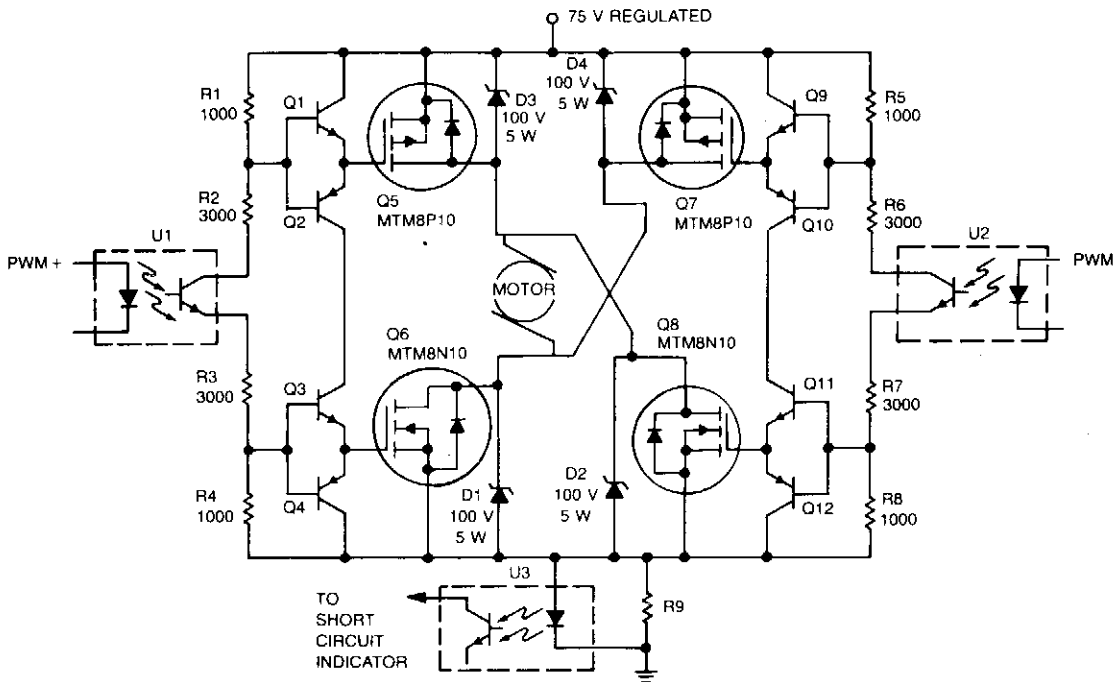


configured as a Wein bridge oscillator. The R1C1, R2C2 feedback networks determine the oscillation frequency, according to the following expression:

$$f_o = \frac{1}{(2\ \pi\sqrt{R_1R_2C_1C_2})}$$

By varying either R1 or R2, the oscillator frequency can be adjusted over a narrow range. The R3/R4 ratio sets the second amplifier's gain to compensate for signal attenuation occurring in the phase shifters. The circuits can be driven from an external source, such as a pulse or square wave, setting the gain of the left-hand amplifier to a level less than that required for oscillation. The RC feedback networks then function as an active filter causing the outputs to be sinusoidal.

SERVO MOTOR DRIVE AMPLIFIER



MOTOROLA

Fig. 60-12

Circuit Notes

Digital ICs and opto-isolators provide the drive for this TMOS servo amplifier, resulting in fewer analog circuits and less drift. Fast and consistent turn-on and turn-off characteristics also enable accurate analog output results directly from the digital signal without the need for analog feedback.

An "H" bridge configuration is employed for the servo amplifier, which obtains complementary PWM inputs from digital control circuits. The PWM inputs are applied via opto-isolators, which keep the digital control logic isolated from the 75 V supply used for the amplifier. A short circuit indicator is provided by opto-isolator U3; if there is a short, the drop across R9 increases to a value sufficient to activate the isolator and send a short indication to the digital control logic.

DC SERVO DRIVE EMPLOYS BIPOLAR CONTROL INPUT

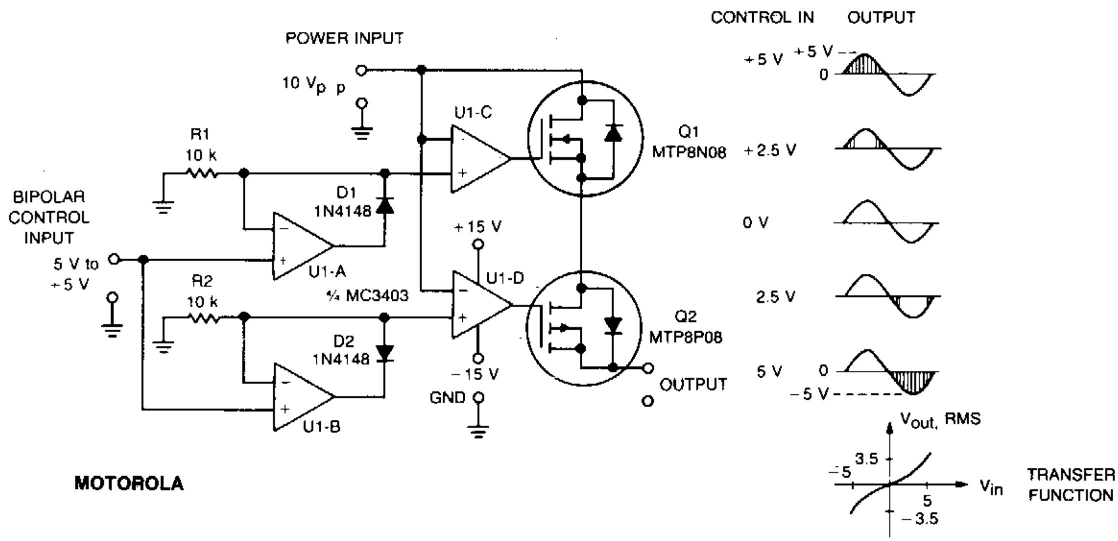


Fig. 60-13

Circuit Notes

This circuit accepts bipolar control inputs of ±5 V and provides a phase-chopped output to a dc load (such as a servo motor) of the same polarity as the input. The rms voltage of the output is closely proportional to the control input voltage.

N-channel and p-channel TMOS devices, Q1 and Q2, are connected in anti-series to form a bidirectional switch through which current can flow in either the forward or reverse direction. Control circuits turn Q1 and Q2 on when they are reverse biased, bypassing their reverse rectifier and increasing circuit efficiency. Each device is allowed to turn off only when forward biased.

The Q1-Q2 switch connects the ac power source to the load when its instantaneous voltage is the same polarity and less than the control voltage. U1a is configured as an ideal positive rectifier whose output follows the control voltage when it is positive, and is zero otherwise. Similarly, U1b is a negative rectifier. U1c turns Q1 on whenever the ac input voltage is lower than the positive rectifier output. For negative control voltages, Q1 is turned on only during the negative half-cycle. For positive control voltages, Q1 is turned on during the end portions of the positive half-cycle. Similarly, U1d turns Q2 on whenever the ac input voltage is higher than the output of the negative rectifier.

400 Hz SERVO AMPLIFIER

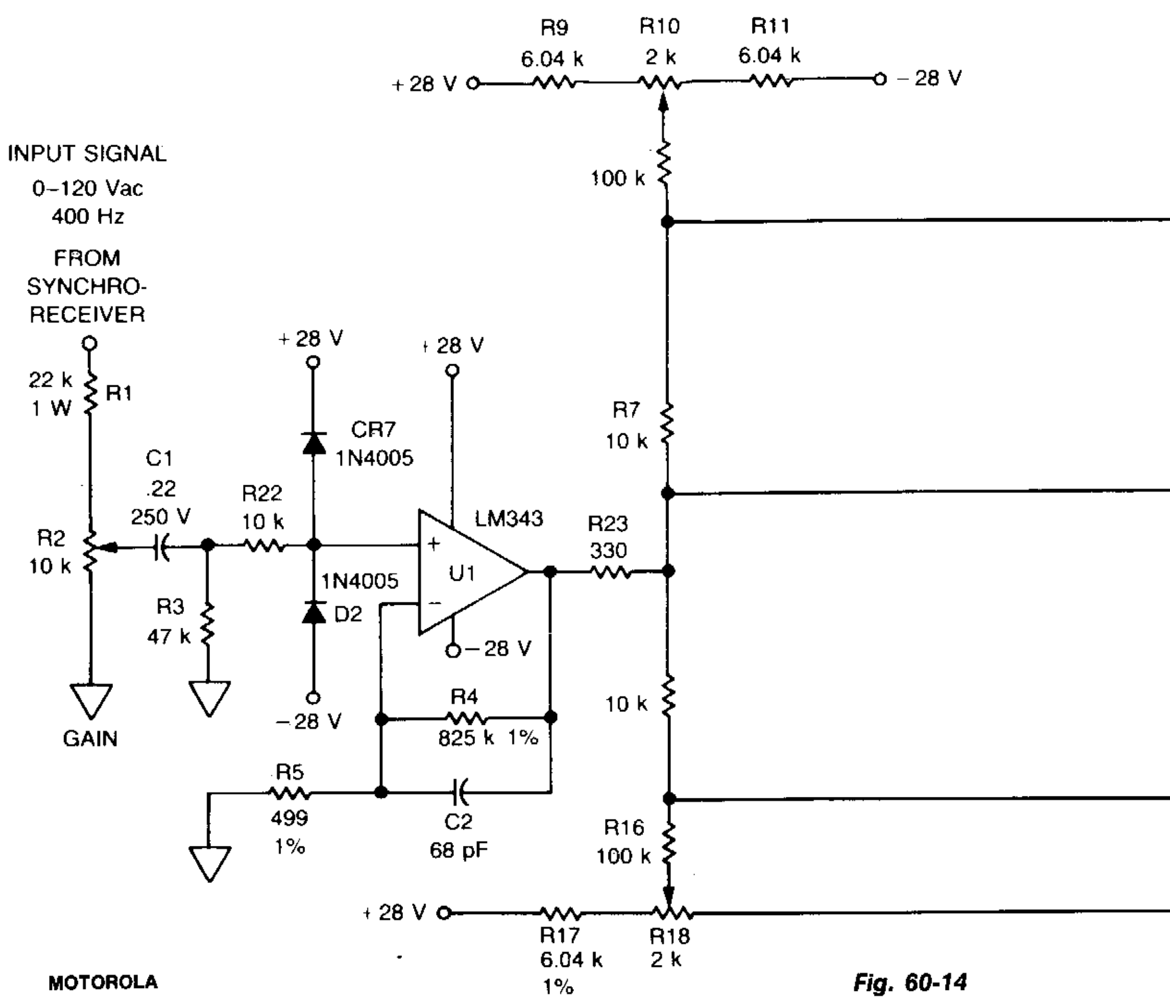
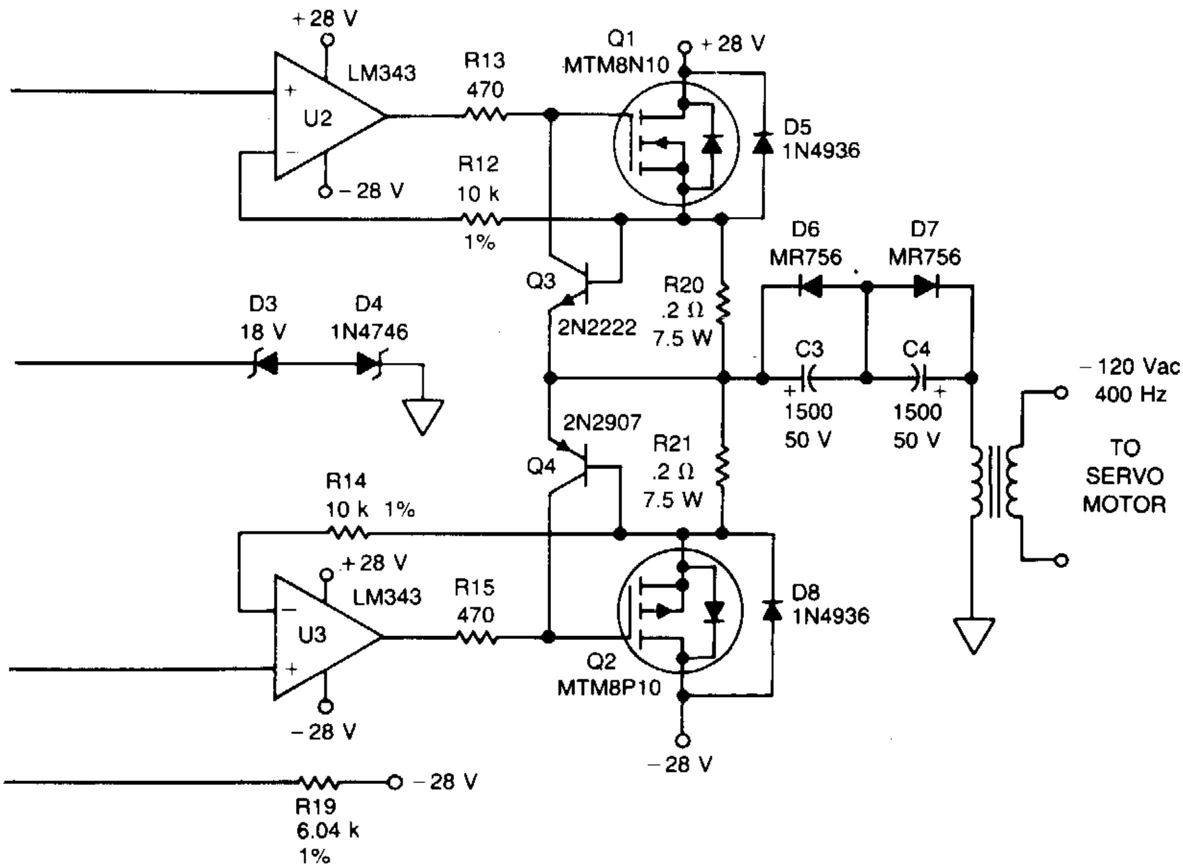


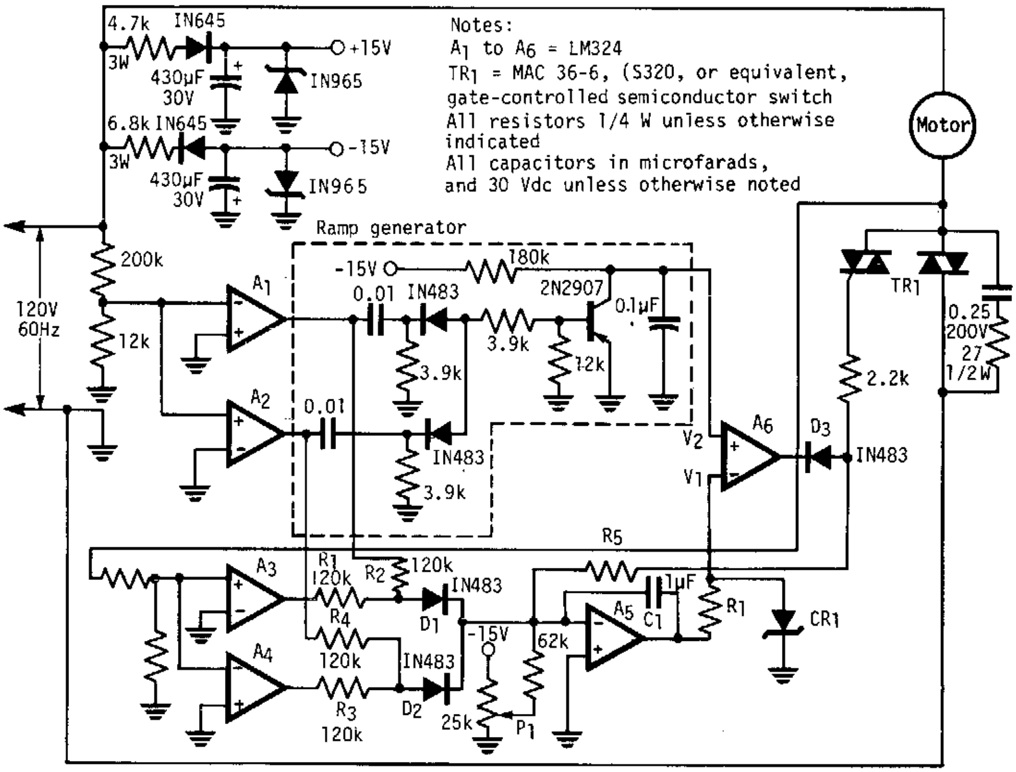
Fig. 60-14

Circuit Notes

The signal from a synchro receiver or a variable resistive cam follower (potentiometer) is boosted by operational amplifier U1, whose output swing is limited by back-to-back zeners D3 and D4. The signal is then applied to operational amplifiers U2 and U3, which drive the gates of Q1 and Q2 respectively. The npn transistor (Q3) is a fast current limiter for the n-channel MTM8N10; a pnp transistor (Q4) performs the same function for the p-channel MTM8P10. Capacitors C3 and C4 eliminate the need for accurate dc offset zeroing. T1 steps up the output voltage to 120 V for the 400 Hz servo motor.


+28 V
LM343
U2
−28 V
R13
470
R12
10 k
1%
Q1
MTM8N10
+28 V
D5
1N4936
D6
MR756
D7
MR756
Q3
2N2222
R20
.2 Ω
7.5 W
D3
18 V
D4
1N4746
C3
1500
50 V
C4
1500
50 V
−120 Vac
400 Hz
TO
SERVO
MOTOR
2N2907
Q4
R21
.2 Ω
7.5 W
R14
10 k 1%
+28 V
LM343
U3
−28 V
R15
470
D8
1N4936
Q2
MTM8P10
−28 V
−28 V
R19
6.04 k
1%

THREE-PHASE POWER-FACTOR CONTROLLER



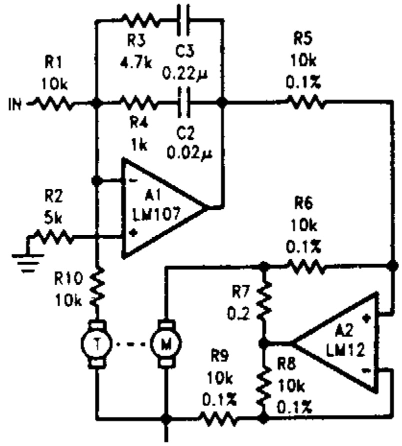
ELECTRONIC ENGINEERING

Fig. 60-15

Circuit Notes

The modified power-factor controller, developed at the Marshall Space Flight Center, employs a phase detector for each of the three phase windings of a delta-connected induction motor. The phase-difference sum is the basis for control. Instabilities of earlier systems are overcome with improved feedback control incorporating a 20Hz bandwidth signal.

MOTOR/TACHOMETER SPEED CONTROL



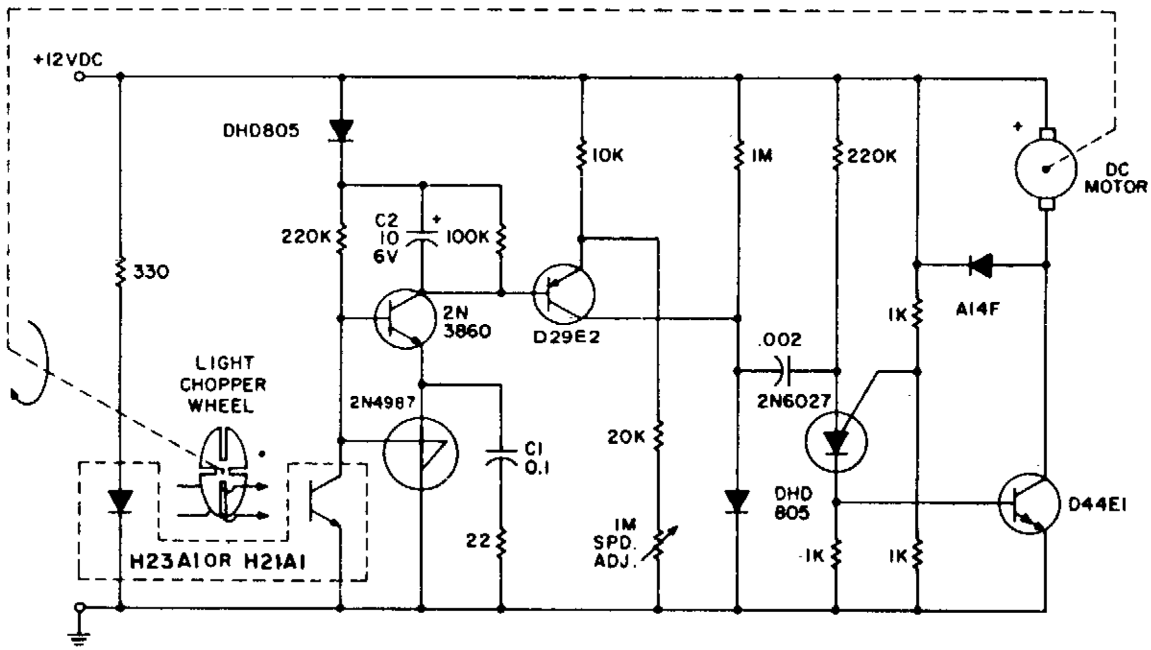
NATIONAL SEMICONDUCTOR CORP.

Fig. 60-16

Circuit Notes

The tachometer, on the same shaft as the dc motor, is simply a generator. It gives a dc output voltage proportional to the speed of the motor. A summing amplifier, A1, controls its output so that the tachometer voltage equals the input voltage, but of opposite sign. With current drive to the motor, phase lag to the tachometer is 90°, before the second order effects come in. Compensation on A1 is designed to give less than 90° phase shift over the range of frequencies where the servo loop goes through unity gain. Should response time be of less concern, a power op amp could be substituted for A1 to drive the motor directly. Lowering break frequencies of the compensation would, of course, be necessary. The circuit could also be used as a position servo. All that is needed is a voltage indicating the sense and magnitude of the motor shaft displacement from a desired position. This error signal is connected to the input, and the servo works to make it zero. The tachometer is still required to develop a phase-correcting rate signal because the error signal lags the motor drive by 180°.

CLOSED LOOP, TACHOMETER FEEDBACK CONTROL



GENERAL ELECTRIC

Fig. 60-17

Circuit Notes

The system utilizes the H21A1 and a chopper disc to provide superior speed regulation when the dynamic characteristics of the motor system and the feedback system are matched to provide stability. The tachometer feedback system illustrated was designed around specific motor/load combinations and may require modification to prevent hunting or oscillation with other combinations. This dc motor control utilizes the optachometer circuit previously shown to control a P.U.T. pulse generator that drives the D44E1 darlington transistor which powers the motor.

61

Multiplier Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ANALOG MULTIPLIER

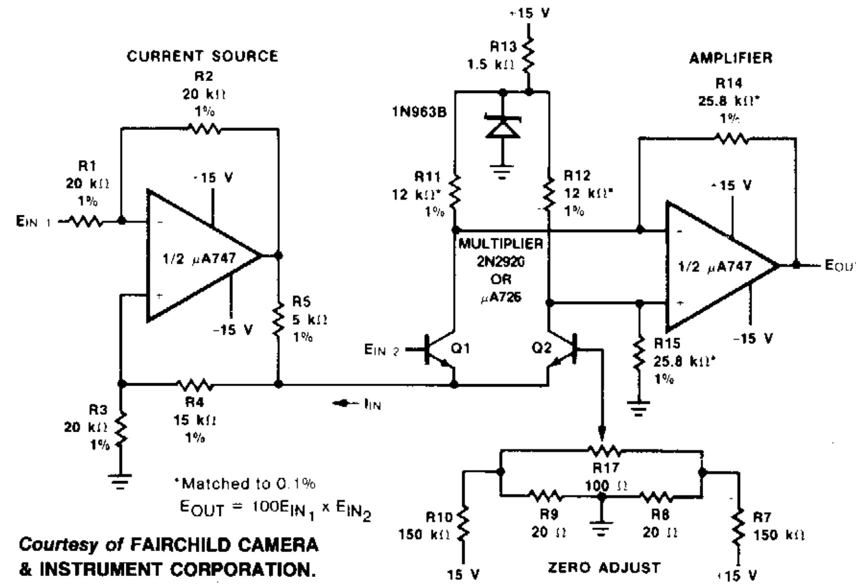
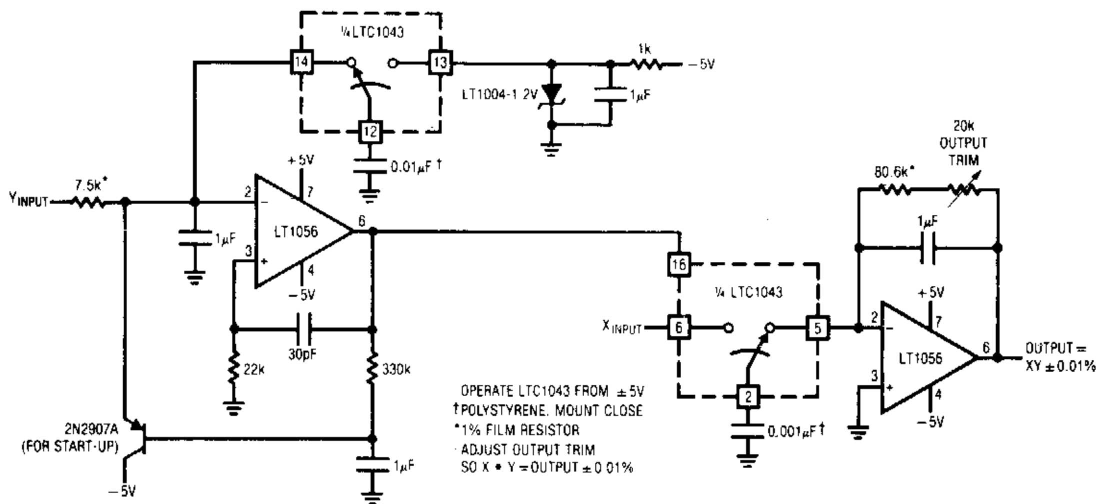


Fig. 61-1

Courtesy of **FAIRCHILD CAMERA & INSTRUMENT CORPORATION.**

0.01% ANALOG MULTIPLIER



LINEAR TECHNOLOGY CORPORATION

Fig. 61-2

Circuit Notes

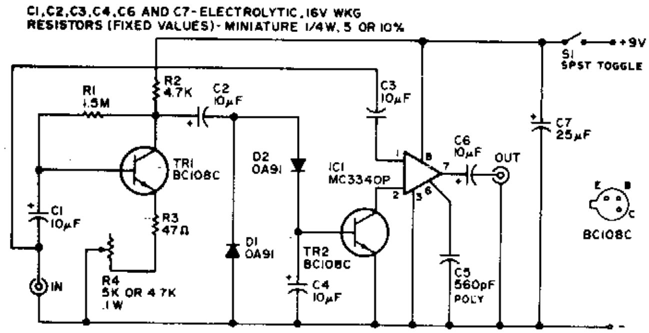
The F → V input frequency is locked to the V → F output because the LTC1043's clock is common to both sections. The F → V's reference is used as one input of the multiplier, while the V → F furnishes the other. To calibrate, short the X and Y inputs to 1.7320 V and trim for a 3-V output.

62

Noise Reduction Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

AUDIO SQUELCH CIRCUIT

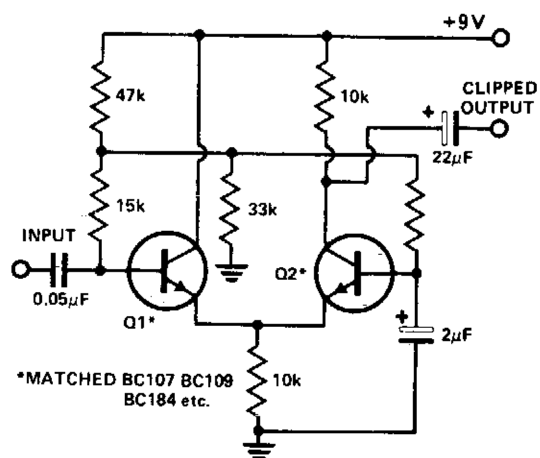


73 MAGAZINE *Fig. 62-1*

Circuit Notes

This simple audio squelch unit suppresses all input signals below a preset threshold.

PRECISE AUDIO CLIPPER

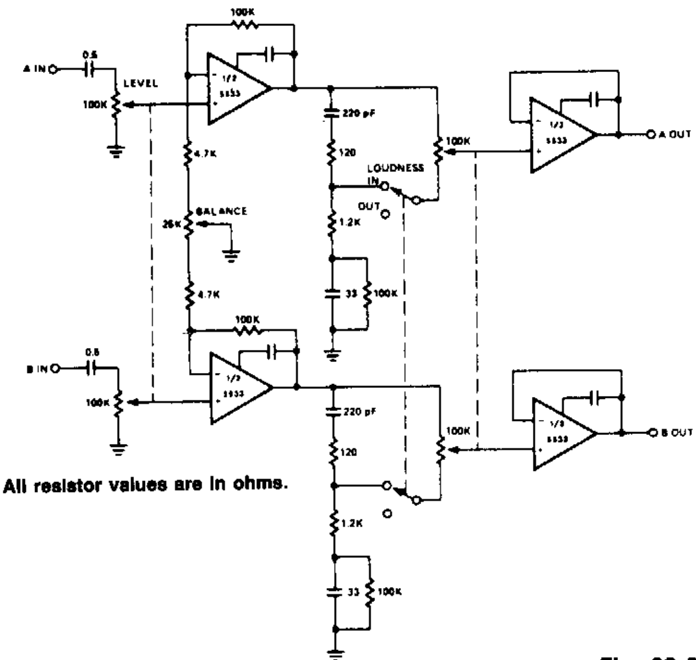


ELECTRONICS INTERNATIONAL TODAY *Fig. 62-2*

Circuit Notes

A differential amplifier makes an excellent audio clipper and can provide precise, symmetrical clipping. The circuit shown commences clipping at an input of 100 mV. The output commences clipping at ±3 V. Matching Q7 and Q2 is necessary for good symmetrical clipping. (If some asymmetry can be tolerated, this need not be done.)

BALANCE AMPLIFIER WITH LOUDNESS CONTROL

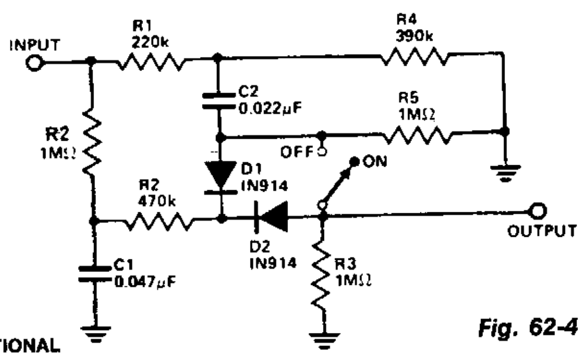


All resistor values are in ohms.

SIGNETICS

Fig. 62-3

NOISE LIMITER



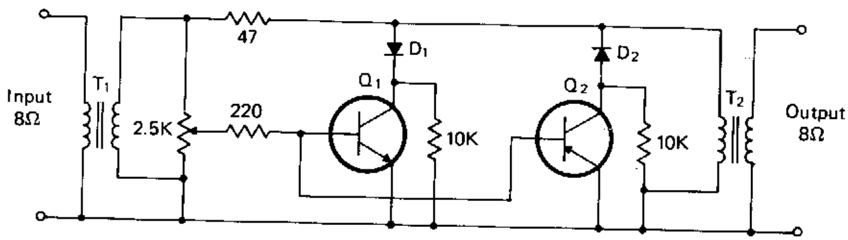
ELECTRONICS TODAY INTERNATIONAL

Fig. 62-4

Circuit Notes

This limiter improves signal-to-noise ratio. It is connected between the detector output and the audio input (if high impedance) or at some relatively high-impedance section between two audio stages—preferably the low level stages. D1 and D2 can be any diode having relatively low forward resistance and very high back resistance. The circuit is excellent for receivers having bandwidths down to 2 or 3 kHz. Increase the value of C1 for receivers having narrower bandwidths.

AUDIO-POWERED NOISE CLIPPER



CQ

Fig. 62-5

Circuit Notes

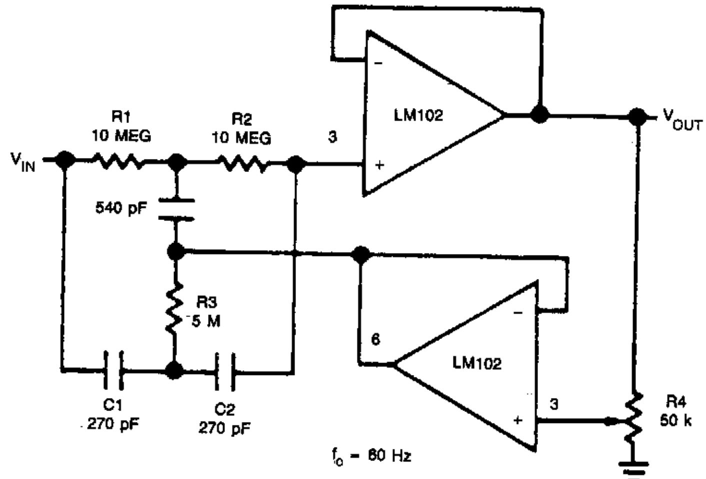
T1 and T2 are 600 to 8 ohm transformers (any transistor radio output transformers with 500 to 4 ohm impedance may be used). Q1 is a 2N2222 npn transistor, and Q2 is a 2N2907 pnp transistor. D1 and D2 1N270 signal diodes (HEP 134 or 135). Two transistors, powered by the audio power contained within the signal, will clip signal peaks which exceed the threshold established by the 2.5 K potentiometer. The diodes isolate the positive and negative clipping circuits represented by the npn and pnp transistors, respectively. A desired audio operating level can be established and the potentiometer needs little or no further adjustment.

63

Notch Filters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ADJUSTABLE Q NOTCH FILTER



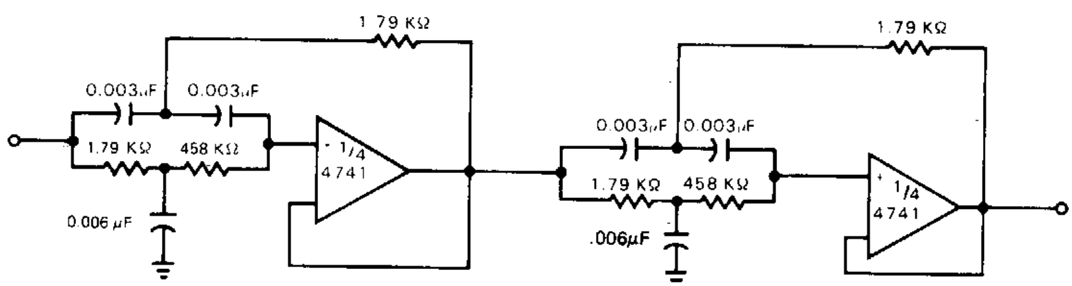
NATIONAL SEMICONDUCTOR CORP.

Fig. 63-1

Circuit Notes

In applications where the rejected signal might deviate slightly from the null on the notch network, it is advantageous to lower the Q of the network. This insures some rejection over a wider range of input frequencies. The figure shows a circuit where the Q may be varied from 0.3 to 50. A fraction of the output is fed back to R3 and C3 by a second voltage follower, and the notch Q is dependent on the amount of signal fed back. A second follower is necessary to drive the twin "T" from a low-resistance source so that the notch frequency and depth will not change with the potentiometer setting.

1800 Hz NOTCH FILTER



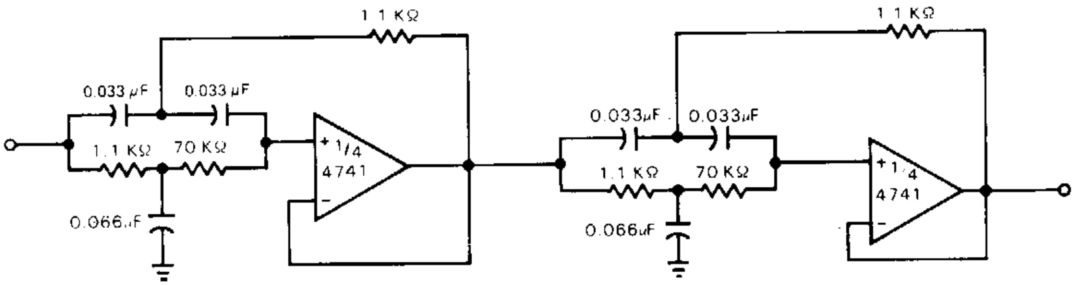
EXAR

Fig. 63-2

Circuit Notes

The circuit produces at least 60 dB of attenuation of the notch frequency.

550 Hz NOTCH FILTER



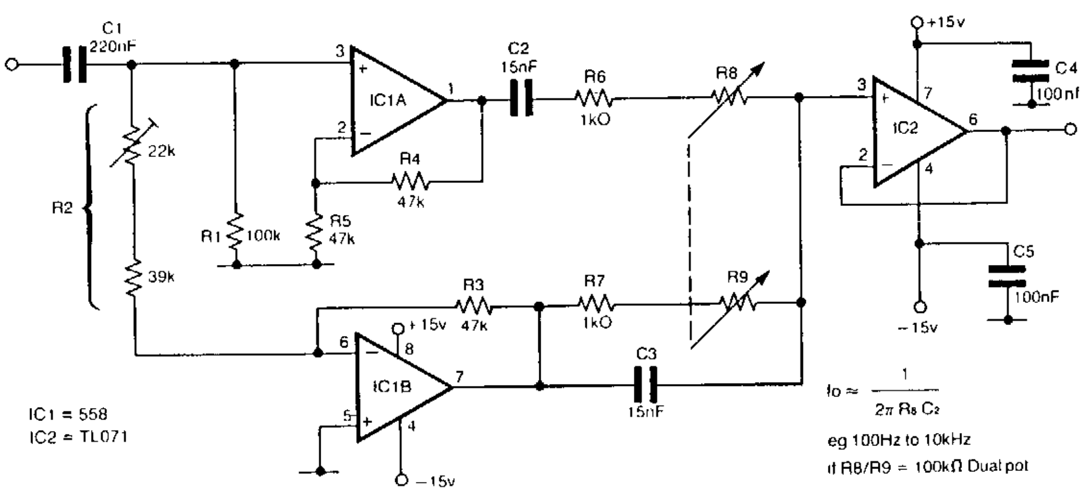
EXAR

Fig. 63-3

Circuit Notes

The circuit produces at least 60 dB of attenuation of the notch frequency.

TUNABLE AUDIO NOTCH FILTER CIRCUIT



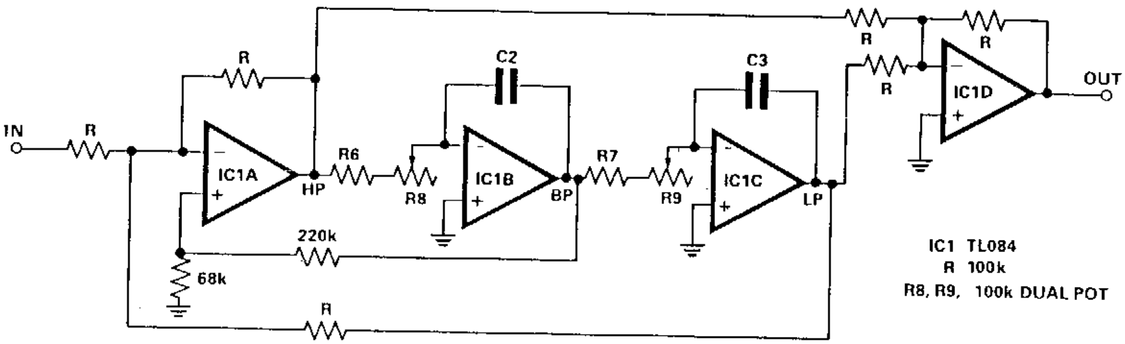
ELECTRONIC ENGINEERING

Fig. 63-4

Circuit Notes

The circuit requires only one dual-ganged potentiometer to tune over a wide range; if necessary over the entire audio range in one sweep. The principle used is that of the Wien bridge, fed from anti-phase inputs. The output should be buffered as shown with a FET input op amp, particularly if a high value pot is used. An op amp with differential outputs (eg., MC1445) may be used in place of the driver ICS; R2 may be made trimmable to optimize the notch.

AUDIO NOTCH FILTER



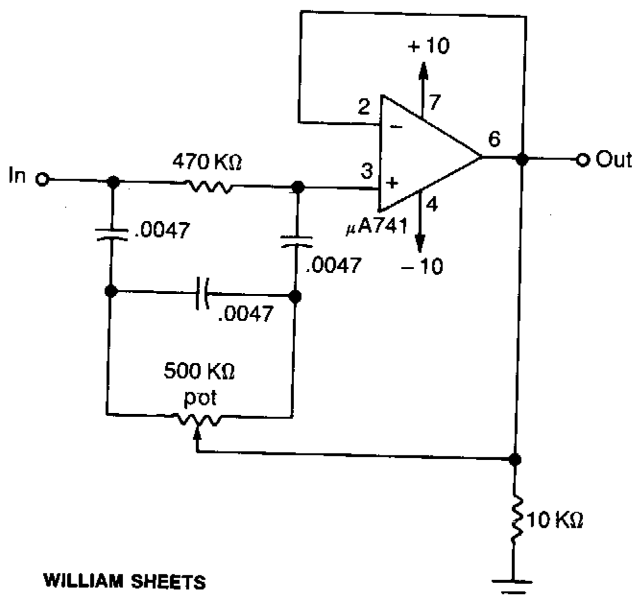
ELECTRONIC ENGINEERING

Fig. 63-5

Circuit Notes

With the circuit shown here the response at one octave off tune is within 10% of the far out response: notch sharpness may be increased or reduced by reducing or increasing respectively the 68 K ohm resistor. Linearity tracking of R8 and R9 has no effect on notch depth. The signals at HP and LP are always in antiphase, notch will always be very deep at the tuned frequency, despite tolerance variations in R6-9 and C2, C3.

TUNABLE NOTCH FILTER USES AN OPERATIONAL AMPLIFIER



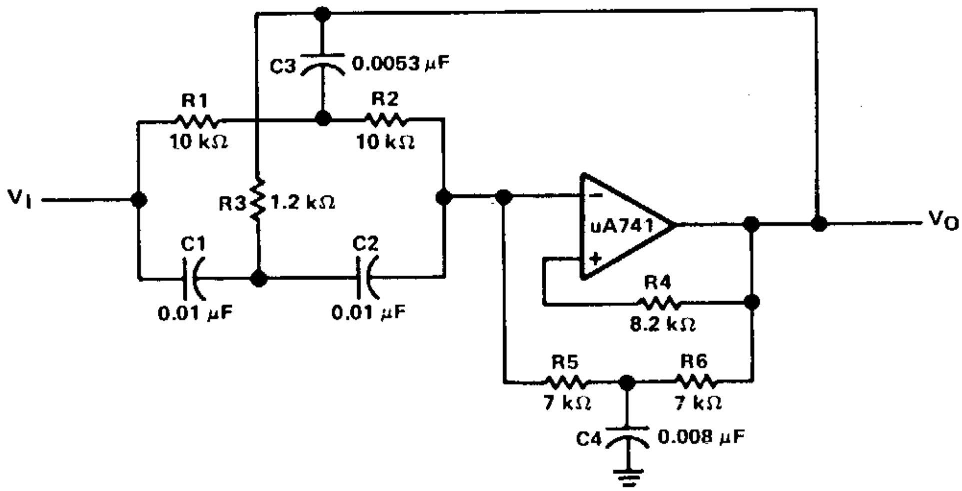
WILLIAM SHEETS

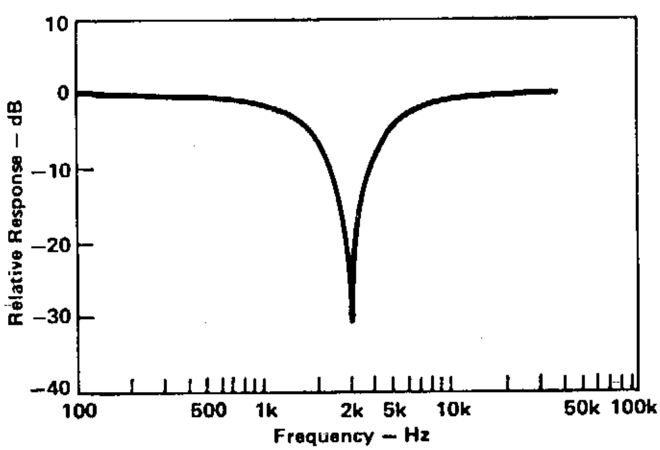
Fig. 63-6

Circuit Notes

This notch filter is useful for tunable band-reject applications in the audio range. The values shown will give a tuning range of about 300-1500 Hz.

ACTIVE BAND-REJECT FILTER



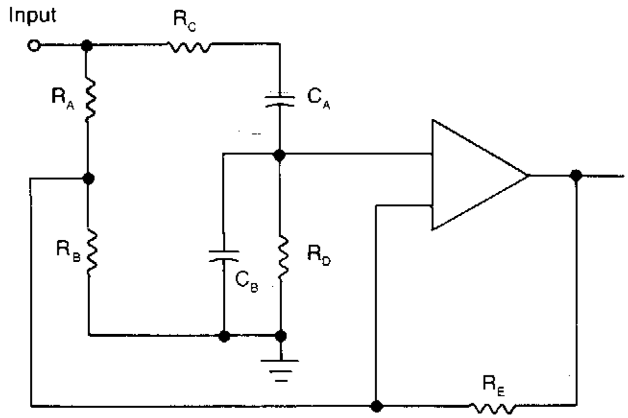


TEXAS INSTRUMENTS *Fig. 63-7*

Circuit Notes

A filter with a band-reject characteristic is frequently referred to as a notch filter. A typical circuit using a μA 741 is the unity-gain configuration for this type of active filter shown. The filter response curve shown is a second-order band-reject filter with a notch frequency of 3 kHz. The resulting Q of this filter is about 23, with a notch depth of −31 dB. Although three passive T networks are used in this application, the operational amplifier has become a sharply tuned low-frequency filter without the use of inductors or large-value capacitors.

WIEN BRIDGE NOTCH FILTER



WILLIAM SHEETS

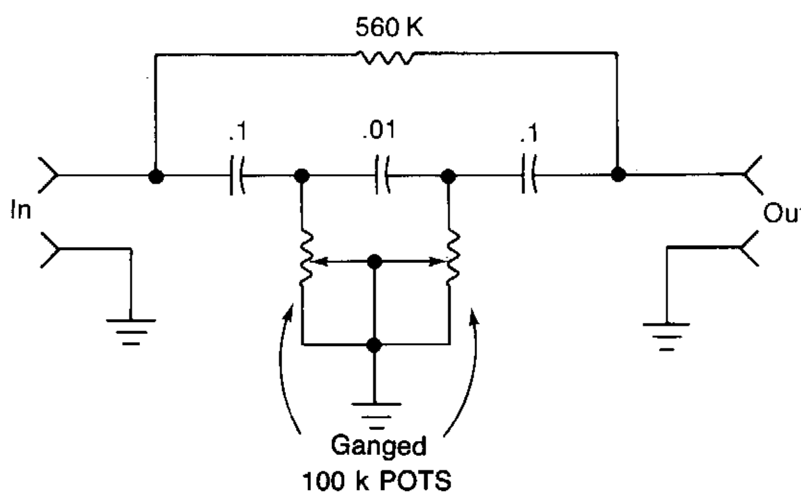
Fig. 63-8

if $R_A = R_B = R_C = R_D = R_E = R$ and $C_A = C_B = C$

$$f_{null} = \frac{1}{6.28\ RC}$$

R megohm
C microfarad
f Hertz

TUNABLE AUDIO FILTER



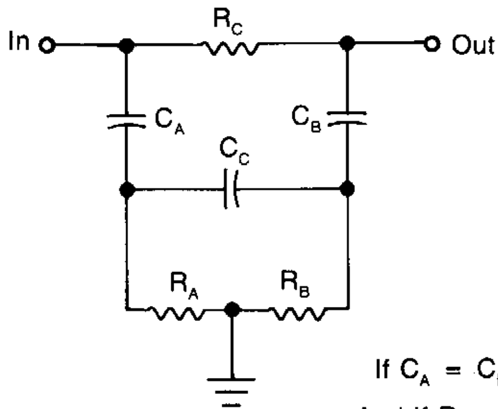
WILLIAM SHEETS.

Fig. 63-9

Circuit Notes

This filter covers the upper part of the audio passband and can be used to eliminate unwanted high frequencies from audio signals.

PASSIVE BRIDGED, DIFFERENTIATOR TUNABLE NOTCH FILTER

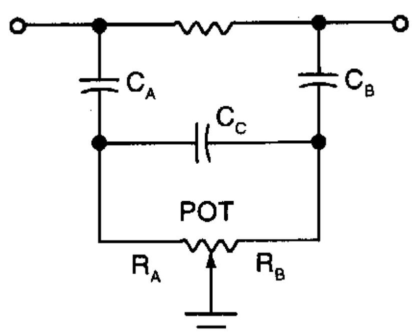


If $C_A = C_B = C_C = C$

And if $R_3 = 6(R_A + R_B)$

Then $\begin{Bmatrix} \text{Notch} \\ \text{freq} \end{Bmatrix} = \dfrac{1}{6.28\ C\sqrt{3R_AR_B}}$

If R_A and R_B is made a potentiometer then the filter can be variable.



WILLIAM SHEETS

Fig. 63-10

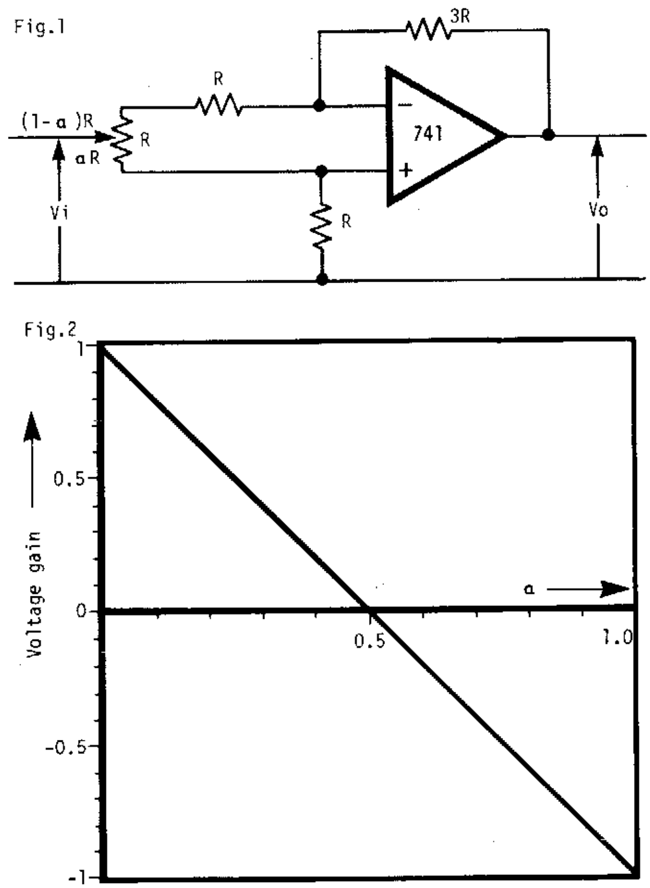
R_A and R_B are sections of potentiometer.

64

Operational Amplifier Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

VARIABLE GAIN AND SIGN OP AMP CIRCUIT



ELECTRONIC ENGINEERING *Fig. 64-1*

Circuit Notes

The gain of the amplifier is smoothly controllable between the limits of +1 to −1. It is adjustable over this range using a single potentiometer. The voltage gain of the arrangement is given by:

$$\frac{V_o}{V_i} = \frac{2(1-2\alpha)}{(1+\alpha)(2-\alpha)}$$

Where α represents the fractional rotation of the potentiometer, R.

SINGLE POTENTIOMETER ADJUSTS OP AMP'S GAIN OVER BIPOLAR RANGE

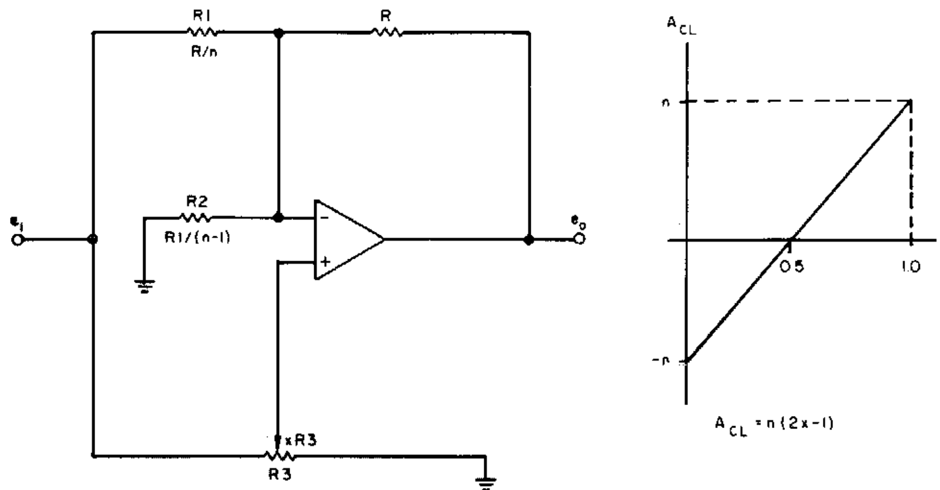


Fig. 64-2

ELECTRONIC DESIGN

Circuit Notes

An op amp's gain level can be adjusted over its full inverting and noninverting gain range. R3 varies the signal applied to both the inverting and noninverting amplifier inputs. When the wiper position (denoted by x) equals zero, the noninverting amplifier input is grounded. This also holds the voltage across R2 at zero, so R2 has no effect on operation. Now only R1 and R carry feedback current, and the amplifier operates at a gain of $-n$. At the other pot extreme, where $x = 1$, the input signal is connected directly to the noninverting input. Since feedback maintains a near-zero voltage between the amplifier inputs, the amplifier's inverting input will also be near the input signal level, thus little voltage is across R1, also now the gain is $+n$. The amplifier should be driven from a low impedance source to minimize source loading error, low offset op amps should be used.

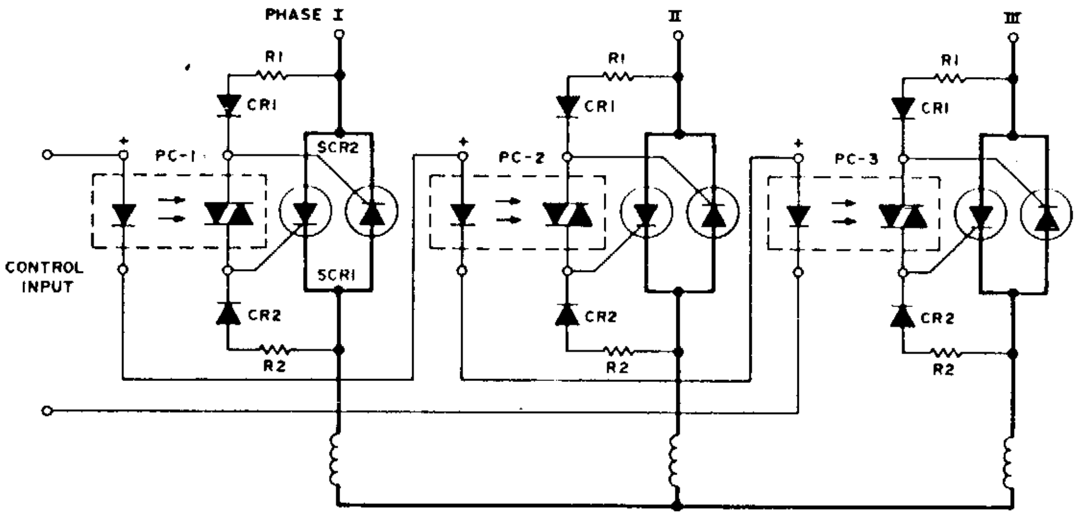
65

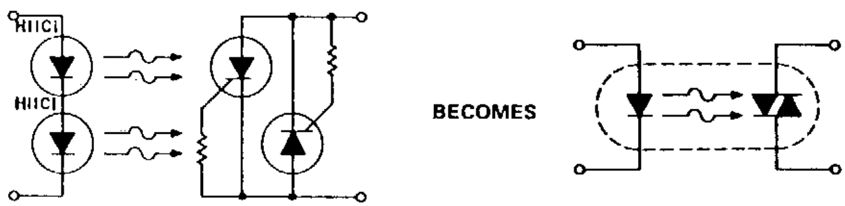
Optically-Coupled Circuits

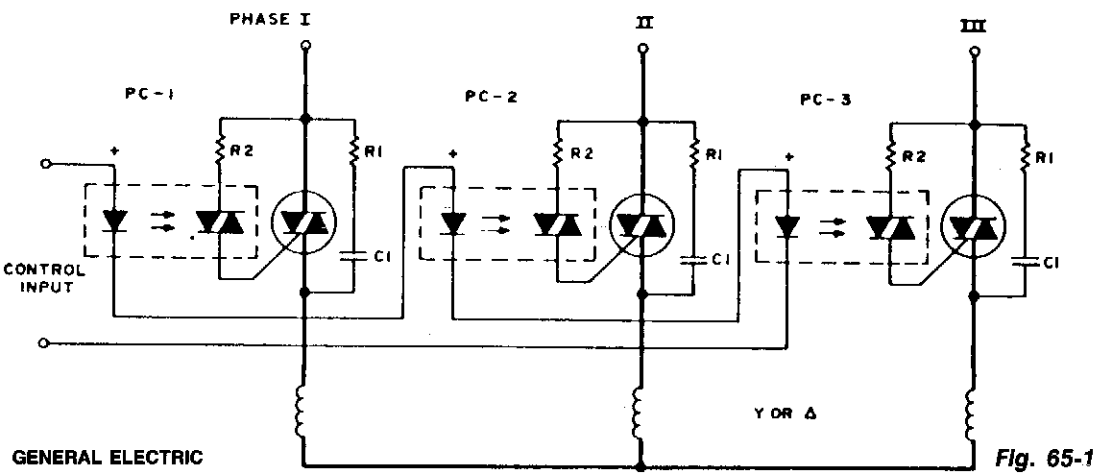
The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Three-Phase Switch for Inductive Load
Integrated Solid State Relay
Stable Optocoupler
Microprocessor Triac Array Driver
DC Linear Coupler
Linear AC Analog Coupler
Simple AC Relay Using Two Photon Couplers
Linear Analog Coupler
High Sensitivity, Normally Open, Two Terminal, Zero Voltage Switching Half-Wave Contact Circuit
High Speed Paper Tape Reader
Digital Transmission Isolator
Isolation and Zero Voltage Switching Logic
Optical Communication System
Linear Optocoupler Circuit for Instrumentation
50 kHz Center Frequency FM Optical Transmitter
Receiver for 50 kHz FM Optical Transmitter

THREE-PHASE SWITCH FOR INDUCTIVE LOAD







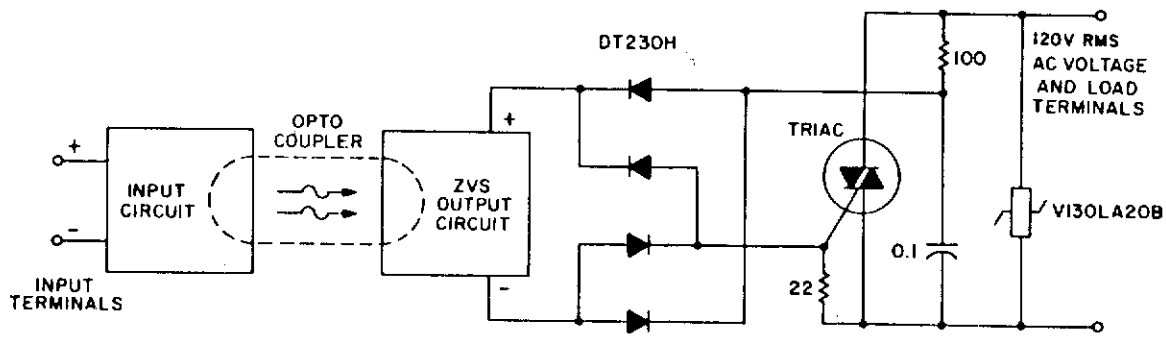
GENERAL ELECTRIC

Fig. 65-1

Circuit Notes

The following are three-phase switches for low voltage. Higher currents can be obtained by using inverse parallel SCRs which would be triggered as shown. To simplify the following schematics and facilitate easy understanding of the principles involved, the following schematic substitution is used (Note the triac driver is of limited use at 3 ϕ voltage levels).

INTEGRATED SOLID STATE RELAY



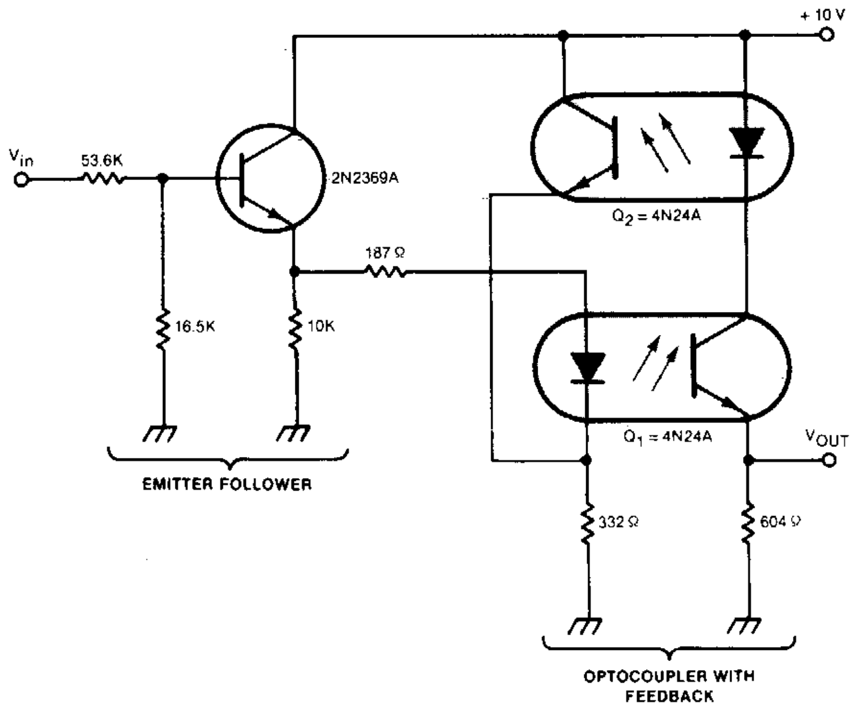
GENERAL ELECTRIC

Fig. 65-2

Circuit Notes

A complete zero-voltage switch solid-state relay contains an input circuit, an output circuit, and the power thyristor. The circuit illustrates a triac power thyristor with snubber circuit and GE-MOV[R] II Varistor transient over-voltage protection. The 22 ohm resistor shunts di/dt currents, passing through the bridge diode capacitances, from the triac gate, while the 100 ohm resistor limits surge and gate currents to safe levels. Although the circuits illustrated are for 120-V rms operation, relays that operate on 220 V require higher voltage ratings on the MOV, rectifier diodes, triac, and pilot SCR. The voltage divider that senses zero crossing must also be selected to minimize power dissipation in the transistor optisolator circuit for 220-V operation.

STABLE OPTOCOUPLER



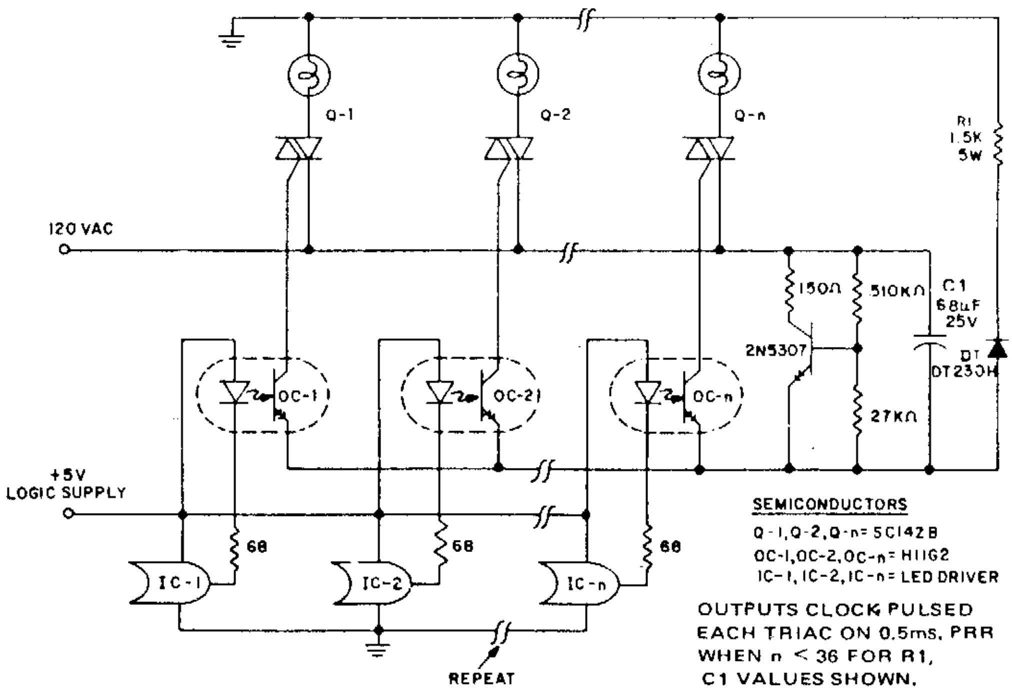
NASA

Fig. 65-3

Circuit Notes

A circuit stabilizes the current-transfer ratio (CTR) of an optically coupled isolator used as a linear transducer. The optocoupler produces a voltage output that is proportional to—but electrically isolated from—the voltage input. However, the output voltage is directly affected by changes in the CTR, and the CTR can change substantially with temperature and current. To a lesser extent the CTR changes with time over the life of the optocoupler. The circuit employs a feedback circuit containing a second optocoupler. The feedback signal tends to oppose changes in the overall CTR.

MICROPROCESSOR TRIAC ARRAY DRIVER



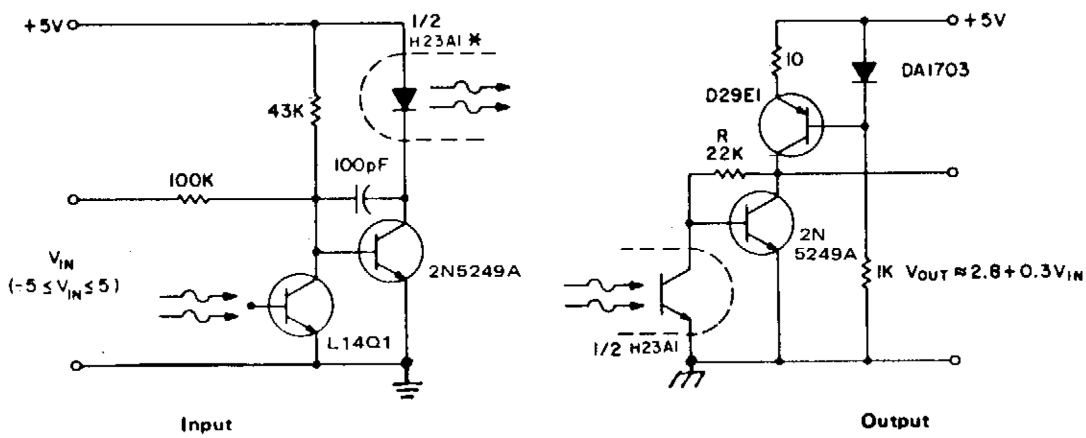
GENERAL ELECTRIC

Fig. 65-4

Circuit Notes

In microprocessor control of multiple loads, the minimum cost per load is critical. A typical application example is a large display involving driving arrays of incandescent lamps. This circuit provides minimal component cost per stage and optocoupler triggering of triac power switches from logic outputs. The minimal component cost is attained by using more complex software in the logic. A darlington output optocoupler provides gate current pulses to the triac, with cost advantages gained from eliminating the current limiting resistor and from the low cost coupler. The trigger current source is a dipped tantalum capacitor, charged from the line via a series resistor with coarse voltage regulation being provided by the darlington signal transistor. The resistor and capacitor are shared by all the darlington-triac pairs and are small in size and cost due to the low duty cycle of pulsing. Coupler IRED current pulses are supplied for the duration of one logic clock pulse (2-10 μsec), at 0.4 to 1 msec intervals, from a LED driver I.C. The pulse timing is derived from the clock waveform when the logic system requires triac conduction. A current limiting resistor is not used, which prevents Miller effect slowdown of the H11G2 switching speed to the extent the triac is supplied insufficient current to trigger. Optodarlington power dissipation is controlled by the low duty cycle and the capacitor supply characteristics.

DC LINEAR COUPLER



LINEAR OPTICAL COUPLER CIRCUIT

*Closely positioned to illuminate L14Q1 and H23A1 Detector, such that $V_{OUT} \cong 2.8V$ at $V_{IN} = 0$.

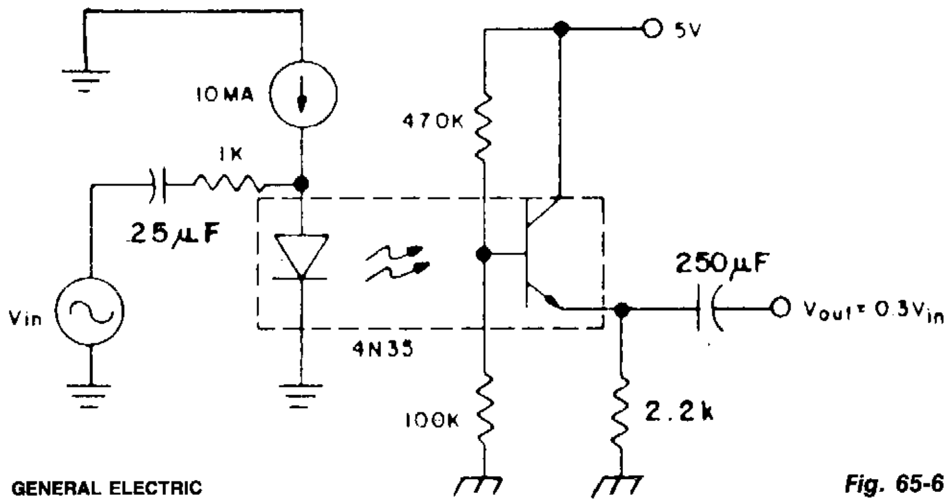
GENERAL ELECTRIC

Fig. 65-5

Circuit Notes

The accuracy of direct linear coupling of analog current signals via an optocoupler is determined by the coupler linearity and its temperature coefficient. Use of an additional coupler for feedback can provide linearity only if the two couplers are perfectly matched and identically biased. These are not practical constraints in most equipment designs and indicate the need for a different design approach. One of the most successful solutions to this problem can be illustrated by using a H23 emitter-detector pair and a L14H4. The H23 detector and L14H4 are placed so both are illuminated by the H23 IRED emitter. Ideally, the circuit is mechanically designed such that the H23 emitter may be positioned to provide $V_{OUT} = 2.8$ V when $V_{IN} = 0$, thereby insuring collector current matching in the detectors. Then all three devices are locked in position relative to each other. Otherwise, R may be adjusted to provide the proper null level, although temperature tracking should prove worse when R is adjusted. Note that the input bias is dependent on power supply voltage, although the output is relatively independent of supply variations. Testing indicated linearity was better than could be resolved, due to alignment motion caused by using plastic tape to lock positions. The concept of feedback control of IRED power output is useful for both information transmission and sensing circuitry.

LINEAR AC ANALOG COUPLER



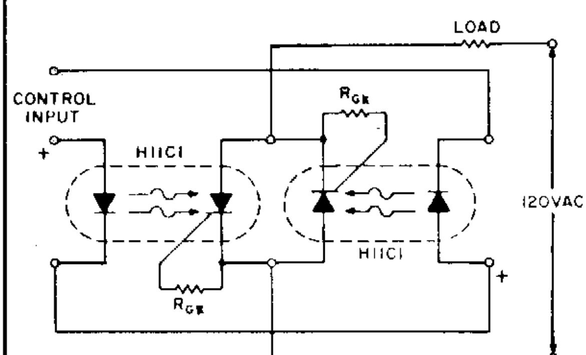
GENERAL ELECTRIC

Fig. 65-6

Circuit Notes

With the coupler biased in the linear region by the 10 mA dc bias on the IRED and the voltage divider on the phototransistor base, photodiode current flows out of the base into the voltage divider, producing an ac voltage proportional to the ac current in the IRED. The transistor is biased as an emitter follower and requires less than 10% of the photodiode current to produce the low impedance ac output across the emitter resistor. Note that the H11AV1 may be substituted for the 4N35 to provide VDE line voltage rated isolation of less than 0.5 pF.

SIMPLE AC RELAY USING TWO PHOTON COUPLERS



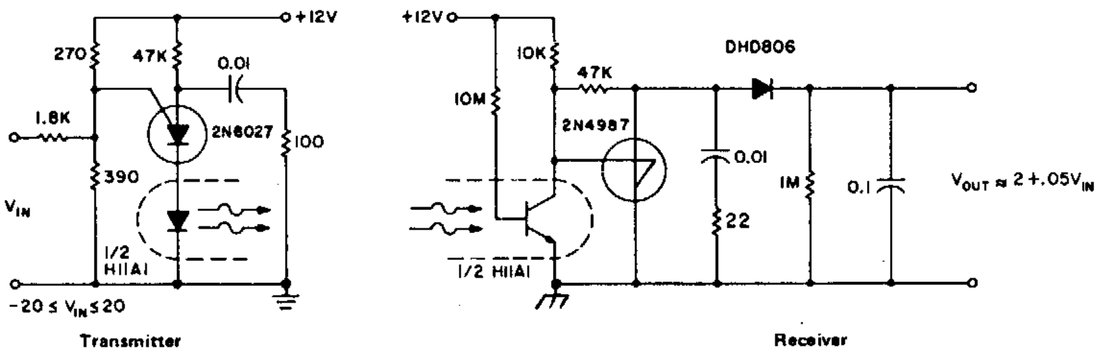
GENERAL ELECTRIC

Fig. 65-7

Circuit Notes

If load current requirements are relatively low (i.e. maximum forward rms current 500 mA), an ac solid state relay can be constructed quite simply by the connection of two H11C optically coupled SCRs in a back-to-back configuration as illustrated.

LINEAR ANALOG COUPLER



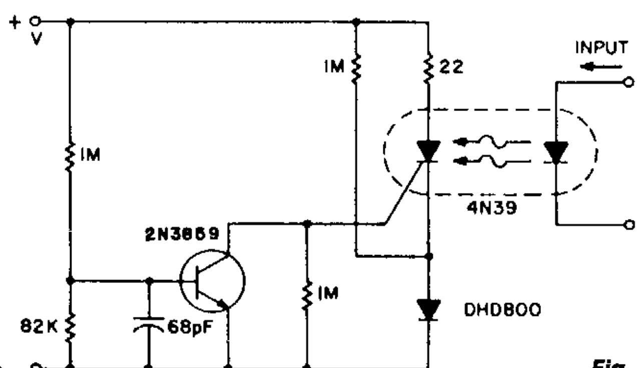
GENERAL ELECTRIC

Fig. 65-8

Circuit Notes

The minimum parts count version of this system provides isolated, linear signal transfer useful at shorter distances or with an optocoupler for linear information transfer. Although the output is low level and cannot be loaded significantly without harming accuracy, a single I.C. operational or instrumentation amplifier can supply both the linear gain and buffering for use with a variety of loads.

HIGH SENSITIVITY, NORMALLY OPEN, TWO TERMINAL, ZERO VOLTAGE SWITCHING, HALF-WAVE CONTACT CIRCUIT



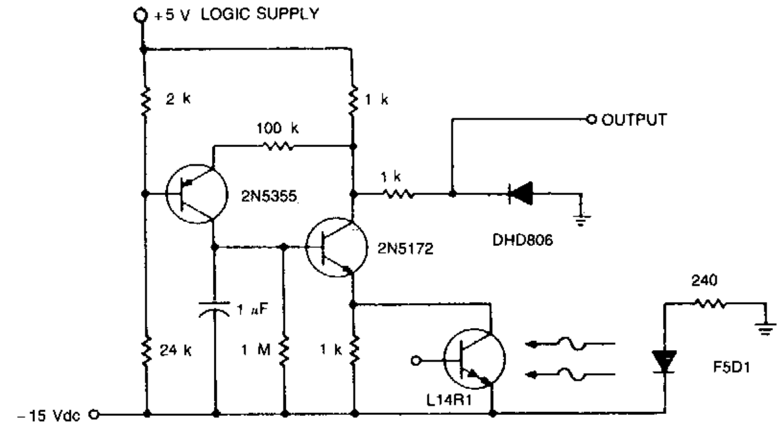
GENERAL ELECTRIC

Fig. 65-9

Circuit Notes

The SCR coupler circuit provides higher sensitivity to input signals as illustrated. This allows the lower cost 4N39 (H11C3) to be used with the > 7 mA drive currents supplied by the input circuit.

HIGH SPEED PAPER TAPE READER



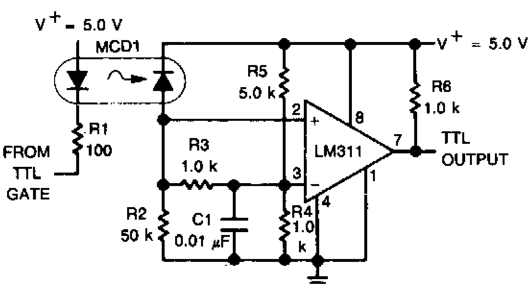
GENERAL ELECTRIC

Fig. 65-10

Circuit Notes

When computer peripheral equipment is interfaced, it is convenient to work with logic signal levels. With a nominal 4 V at the output dropping to −0.6 V on illumination, this circuit reflects the requirements of a high-speed, paper tape optical reader system. The circuit operates at rates of up to 1000 bits per second. It will also operate at tape translucency such that 50% of the incident light is transmitted to the sensor, and provide a fixed threshold signal to the logic circuit, all at low cost. Several circuit tricks are required. Photodarlington speed is enhanced by cascode constant voltage biasing. The output threshold and tape translucency requirements are provided for by sensing the output voltage and operating to 2000 bits per second at ambient light levels equal to signal levels.

DIGITAL TRANSMISSION ISOLATOR



National Semiconductor Corp.

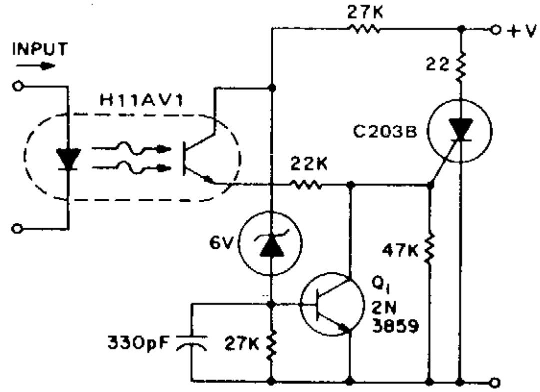
Fig. 65-11

Circuit Notes

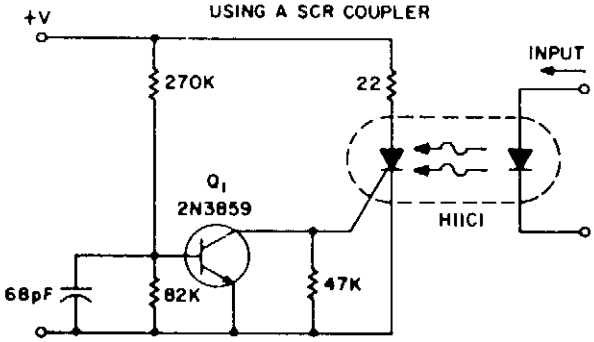
An optoelectronics device is used to couple a digital (TTL) signal to another system. The photodiode in the optocoupler drives an LM311 set up to produce a TTL compatible output. It is useful where grounds are not able to be connected for any reason.

ISOLATION AND ZERO VOLTAGE SWITCHING LOGIC

USING A TRANSISTOR OPTOCOUPLER



USING A SCR COUPLER



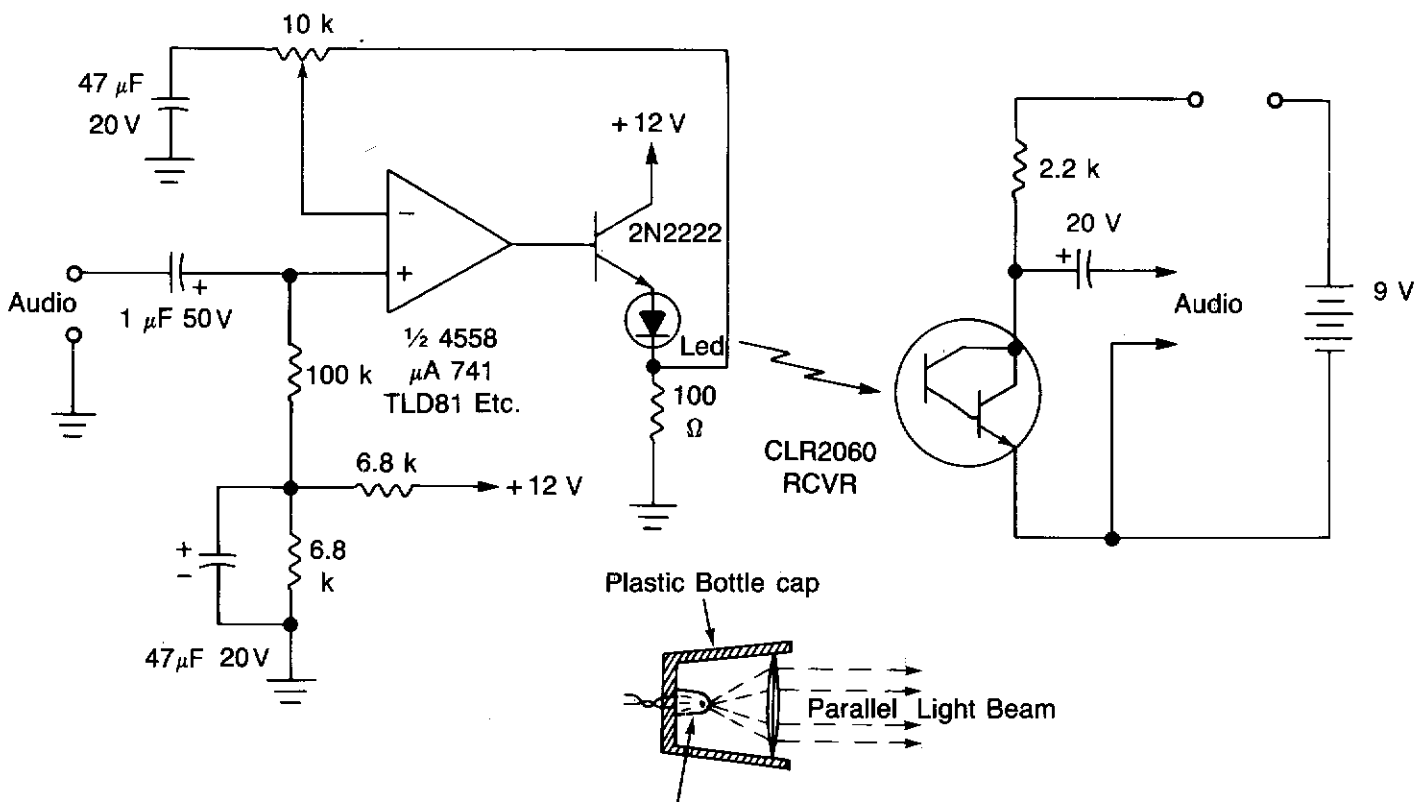
NORMALLY OPEN, TWO TERMINAL, ZERO VOLTAGE SWITCHING HALF WAVE CONTACT CIRCUITS

GENERAL ELECTRIC *Fig. 65-12*

Circuit Notes

These two simple circuits provide zero voltage switching. They can be used with full wave bridges or in antiparallel to provide full wave control and are normally used to trigger power thyristors. If an input signal is present during the time the ac voltage is between 0 to 7 V, the SCR will turn on. But, if the ac voltage has risen above this range and the input signal is then applied, the transistor, Q1, will be biased to the "on" state and will hold the SCR and, consequently, the relay "off" until the next zero crossing.

OPTICAL COMMUNICATION SYSTEM



WILLIAM SHEETS

Fig. 65-13

Circuit Notes

The circuit will modulate the light from the LED using a crystal microphone or a loudspeaker output. To obtain the maximum range, the optical system must be efficient (see example). Either a convex lens or a concave mirror can be used to convert the LED output into a parallel beam. The received light is concentrated onto a sensitive photodarlington transistor. At short range the signal across the load resistor is adequate to drive a crystal earpiece, for longer range an amplifier and a loudspeaker are needed.

50 kHz CENTER FREQUENCY FM OPTICAL TRANSMITTER

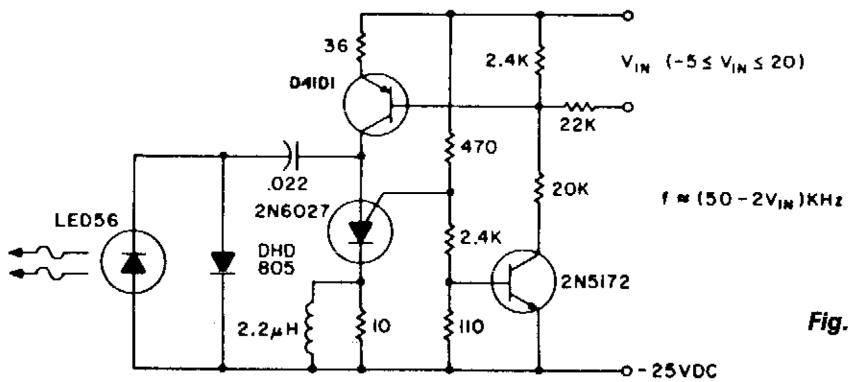


Fig. 65-14

GENERAL ELECTRIC

Circuit Notes

The pulse repetition rate is relatively insensitive to temperature, and power supply voltage and is a linear function of V_{IN}, the modulating voltage. Useful information transfer was obtained in free air ranges of 12 feet (≈ 4m). Lenses or reflectors at the light emitter and detector increases range and minimizes stray light noise effects. Greater range can also be obtained by using a higher power output IRED such as the F5D1 in combination with the L14P2 phototransistor. Average power consumption of the transmitter circuit is less than 3 watts.

LINEAR OPTOCOUPLER CIRCUIT FOR INSTRUMENTATION

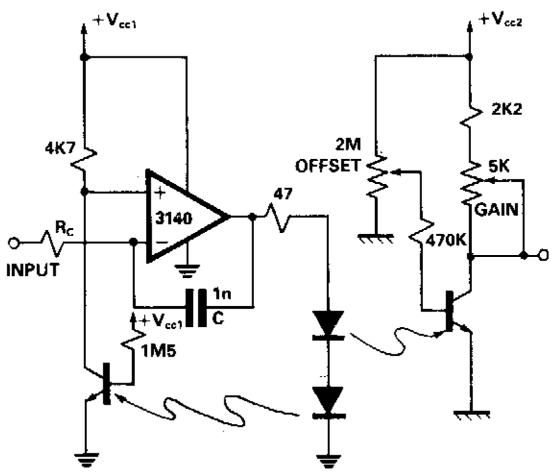


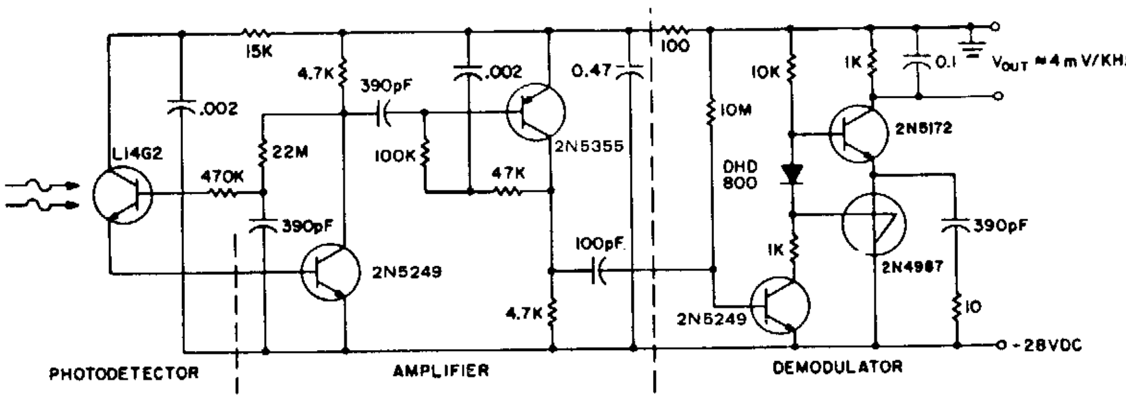
Fig. 65-15

Circuit Notes

A dual optocoupler is used in a configuration which has the same current throughout as the LEDs. Assuming similar optocoupler features the output voltage must be equal to the non-inverting input voltage. Since the op amp is within a closed loop the output voltage becomes equal to the input voltage. R_c and C perform as a compensation network to prevent oscillations.

ELECTRONIC ENGINEERING

RECEIVER FOR 50 kHz FM OPTICAL TRANSMITTER



GENERAL ELECTRIC

Fig. 65-16

Circuit Notes

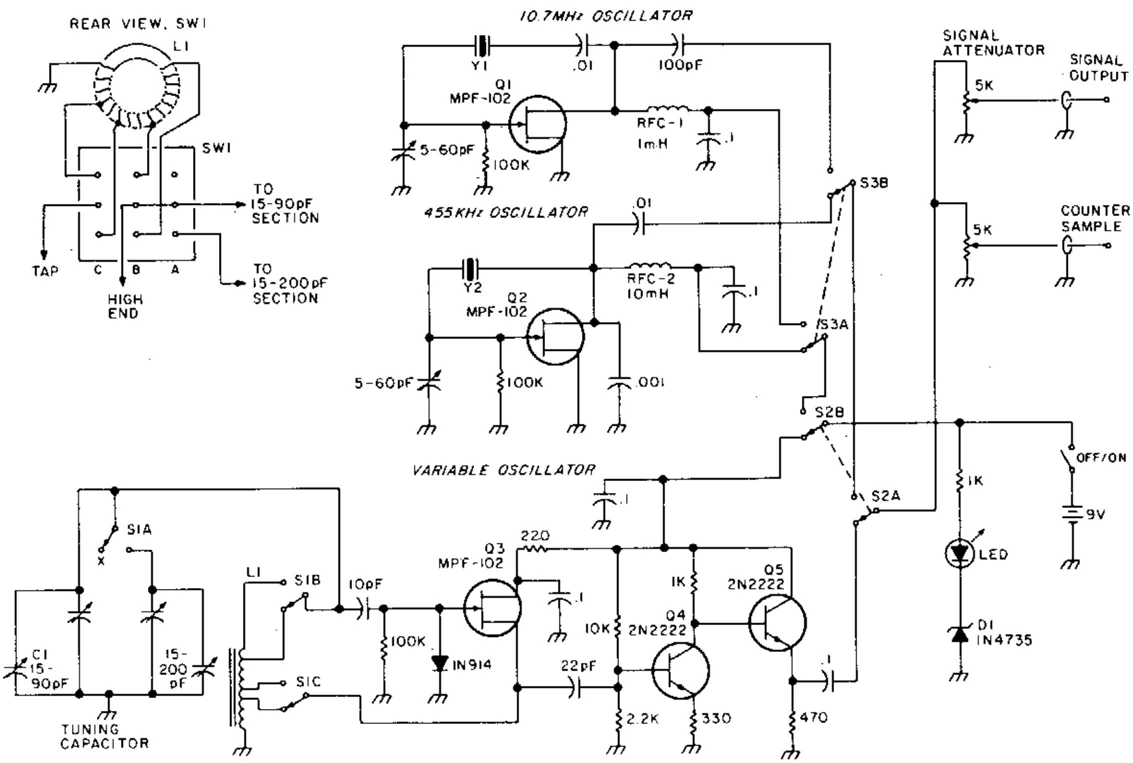
For maximum range, the receiver must be designed in the same manner as a radio receiver front end, since the received signals will be similar in both frequency component and in amplitude of the photodiode current. The major constraint on the receiver performance is signal to noise ratio, followed by e.m. shielding, stability, bias points, parts layout, etc. These become significant details in the final design. This receiver circuit consists of a L14G2 detector, two stages of gain, and a FM demodulator which is the tachometer circuit, modified to operate up to 100 kHz. Better sensitivity can be obtained using more stages of stabilized gain with AGC, lower cost and sensitivity may be obtained by using an H23A1 emitter-detector pair and/or by eliminating amplifier stages. For some applications, additional filtering of the output voltage may be desired.

66

Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RF-GENIE



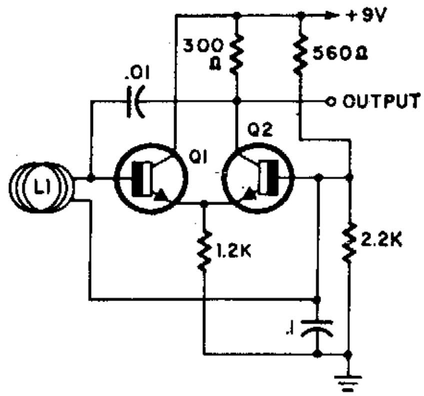
73 MAGAZINE

Fig. 66-1

Circuit Notes

A variable oscillator covers 3.2 to 22 MHz in two bands—providing coverage of 80 through 15 meters plus most crystal-filter frequencies. Optional 455 kHz and 10.7 MHz crystal oscillators can be switched on-line for precise *if* alignment. Generator output is on the order of 4 volts p-p into a 500 ohm load. A simple voltage-divider attenuator controls the generator's output level, and a second output provides sufficient drive for an external frequency counter.

EMITTER-COUPLED BIG LOOP OSCILLATOR

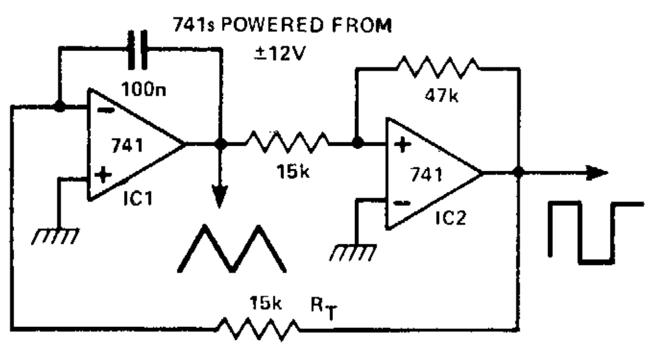


RADIO-ELECTRONICS *Fig. 66-2*

Circuit Notes

L1 is a loop of 10 to 20 turns of insulated wire with a diameter anywhere between 4″ to 4′. Oscillator frequency (7 to 30 MHz) shifts substantially when a person comes near or into the loop. This oscillator together with a resonant detector might make a very good anti-personnel alarm. Transistors are 2N2926 or equivalent.

SIMPLE TRIANGLE SQUARE WAVE OSCILLATOR

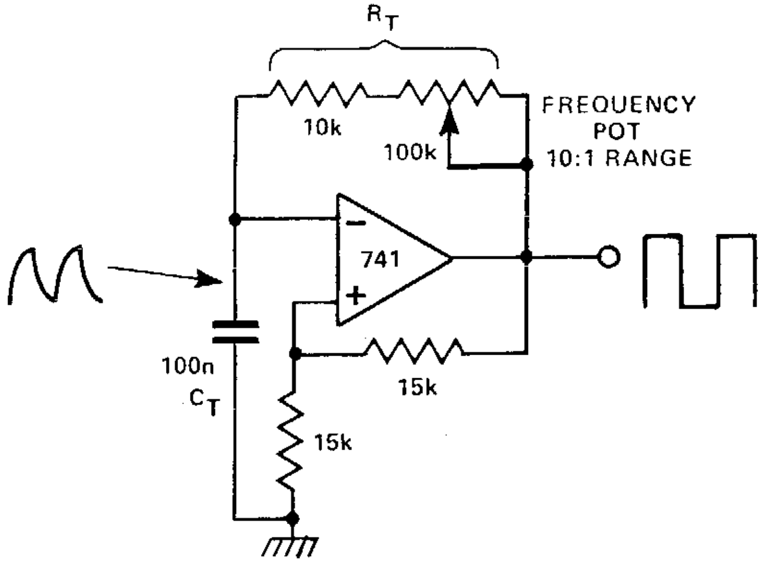


ELECTRONICS TODAY INTERNATIONAL *Fig. 66-3*

Circuit Notes

This circuit generates simultaneously, a triangle and a square waveform. It is self starting and has no latch up problems. IC1 is an integrator with a slew rate determined by CT and RT and IC2 is a Schmitt trigger. The output of IC1 ramps up and down between the hysteresis levels of the Schmitt, the output of which drives the integrator. By making RT variable, it is possible to alter the operating frequency over a 100 to 1 range. Three resistors, one capacitor, and a dual op amp is all that is needed to make a versatile triangle and square wave oscillator with a possible frequency range of 0.1 Hz to 100 kHz.

OSCILLATOR ADJUSTABLE OVER 10:1 RANGE



ELECTRONICS TODAY INTERNATIONAL

Fig. 66-4

Circuit Notes

In this circuit, there are two feedback paths around an op-amp. One is positive dc feedback which forms a Schmitt trigger. The other is a CR timing network. Imagine that the output voltage is +10 V. The voltage at the noninverting terminal is +15 V. The voltage at the inverting terminal is a rising voltage with a time constant of $C_T R_T$. When this voltage exceeds +5 V, the op amp's output will go low and the Schmitt trigger action will make it snap into its negative state. Now the output is −10 V and the voltage at the inverting terminal falls with the time constant as before. By changing this time constant with a variable resistor, a variable frequency oscillation may be produced.

ONE SECOND, 1 kHz OSCILLATOR

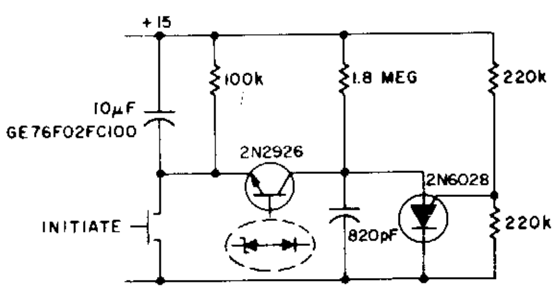


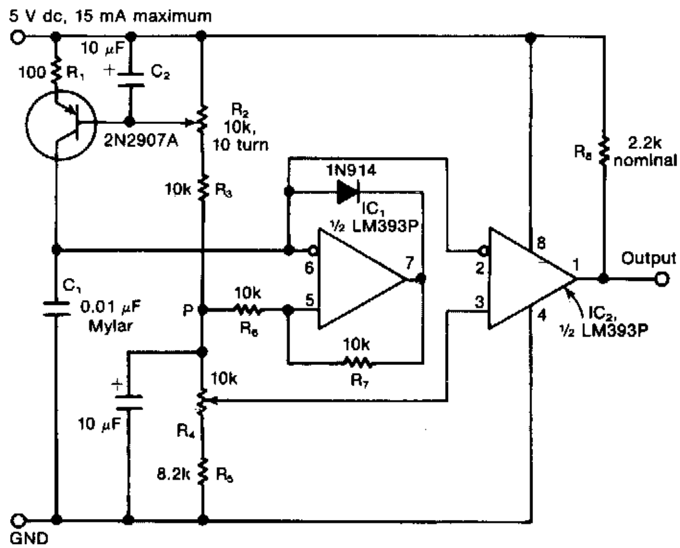
Fig. 66-5

GENERAL ELECTRIC

Circuit Notes

This circuit operates as an oscillator and a timer. The 2N6028 is normally on due to excess holding current through the 100 k resistor. When the switch is momentarily closed, the 10 μF capacitor is charged to a full 15 volts and 2N2926 starts oscillating (1.8 M and 820 pF). The circuit latches when 2N2926 zener breaks down again.

SINGLE CONTROL FOUR-DECADE VARIABLE OSCILLATOR



ELECTRONIC DESIGN

Fig. 66-6

Circuit Notes

The circuit consists of a variable current source that charges a capacitor, which is rapidly discharged by a Schmitt-trigger comparator. The sawtooth waveform thus produced is fed to another comparator, one with a variable switching level. The output from the second comparator is a pulse train with an independently adjustable frequency and duty cycle. The variable-frequency ramp generator consists of capacitor C1, which is charged by a variable and nonlinear current source. The latter comprises a 2N2907A pnp transistor, plus resistor R1 and the potentiometer R2. Capacitor C2 eliminates any ripple or noise at the base of the transistor that might cause frequency jitter at the output.

TUNABLE FREQUENCY OSCILLATORS

FREQUENCY RANGE
40 Hz to 65 kHz

OUTPUT PULSE

Rise time ~ 200 nsec.
Pulse width ~ 10 μsec.
Recovery time < 200 nsec.

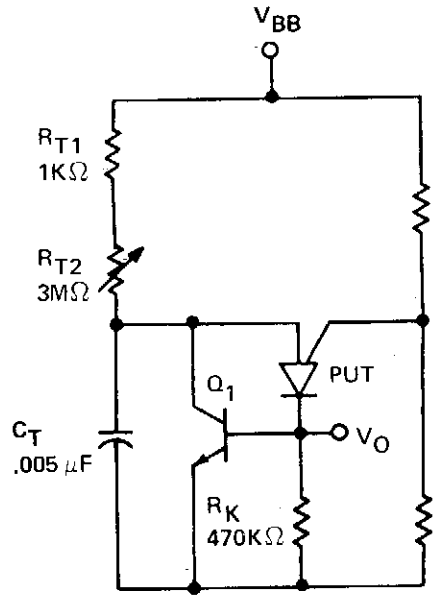


Fig. 66-7A

FREQUENCY RANGE
40 Hz to 40 kHz

OUTPUT PULSE

Width ~ 5 μsec.

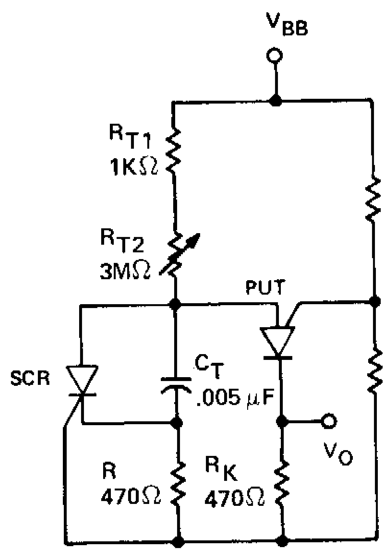


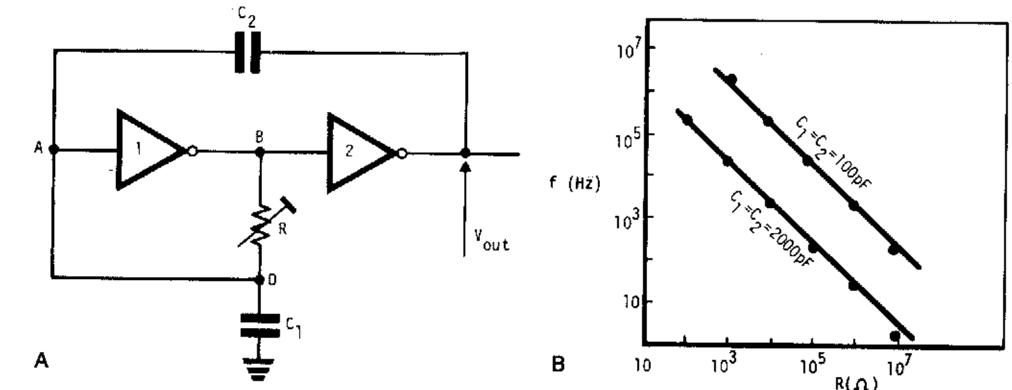
Fig. 66-7B

UNITRODE CORPORATION

Circuit Notes

The variable oscillator circuit includes active elements for discharging the timing capacitor C_T shown in Fig. 66-7A. A second method is given as in Fig. 66-7B.

RESISTANCE CONTROLLED DIGITAL OSCILLATOR



ELECTRONIC ENGINEERING

Fig. 66-8

Circuit Notes

This very simple, low cost oscillator, is built with two CMOS buffer inverters, two capacitors and a variable resistance. The circuit can work with voltages ranging from 4 V up to 18 V. If C1 = C2, the frequency of oscillation, (ignoring the output and input impedance) is given by:

$$f = \frac{1}{4\pi\sqrt{2RC}}$$

The graph in Fig. B shows how the output frequency varies with resistance when C1 = C2 = 100 pF and C1 = C2 = 2000 pF.

CASSETTE BIAS OSCILLATOR

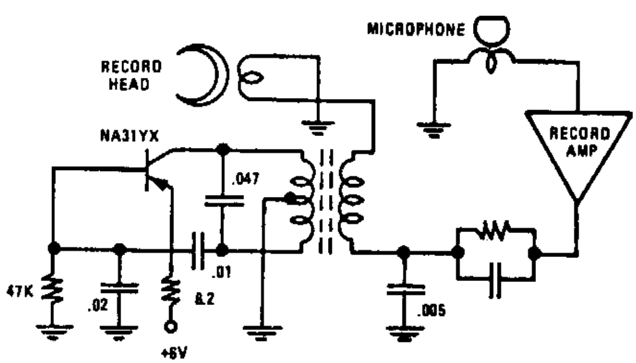
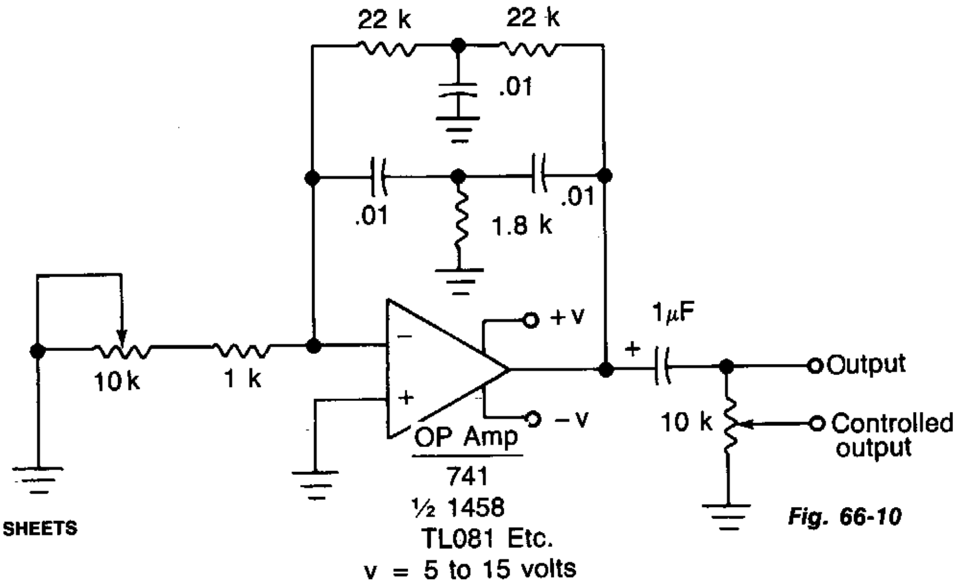


Fig. 66-9

NATIONAL SEMICONDUCTOR

1 kHz OSCILLATOR



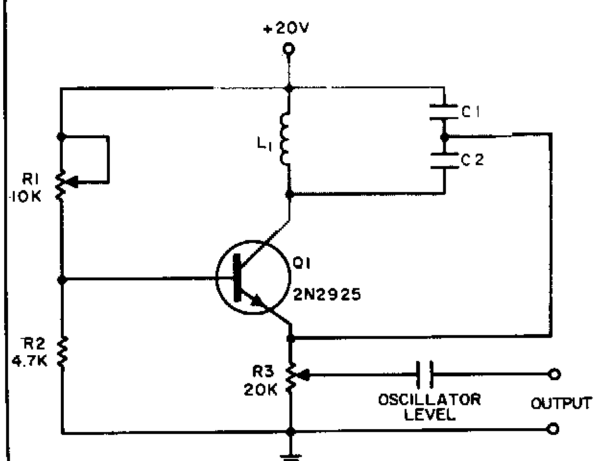
WILLIAM SHEETS

Fig. 66-10

Circuit Notes

If fine output control is desired, add the 10 K potentiometer. When the oscillator is connected to a dc circuit then connect a dc blocking capacitor in series with the potentiometer's wiper arm.

INEXPENSIVE OSCILLATOR IS TEMPERATURE STABLE



ELECTRONIC DESIGN

Fig. 66-11

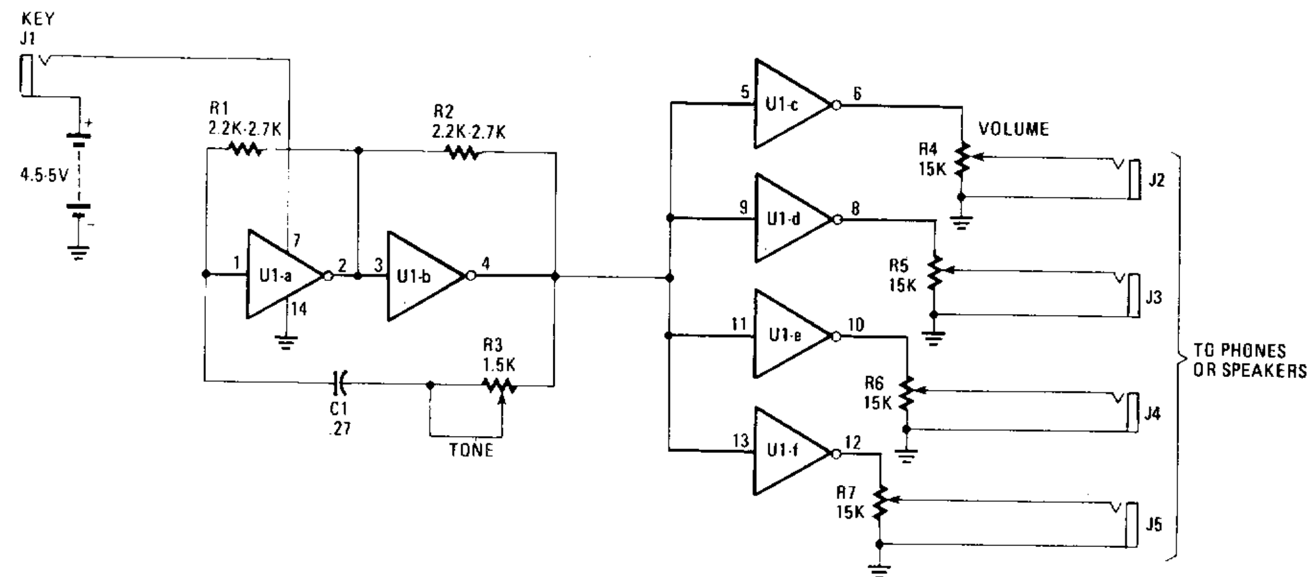
Circuit Notes

The Colpitts sinusoidal oscillator provides stable output amplitude and frequency from 0°F to +150°F. In addition, output amplitude is large and harmonic distortion is low. Oscillation is sustained by feedback from the collector tank circuit to the emitter. The oscillator's frequency is determined by:

$$f = \frac{1}{2\pi \sqrt{\frac{L1C1C2}{C1 + C2}}}$$

Potentiometer R3 is an output level control. Control R1 may be used to adjust base bias for maximum-amplitude output. The circuit was operated at 50 kHz with L1 = 10mH, C1 = 3500 pF, and C2 = 1500 pF.

CODE PRACTICE OSCILLATOR



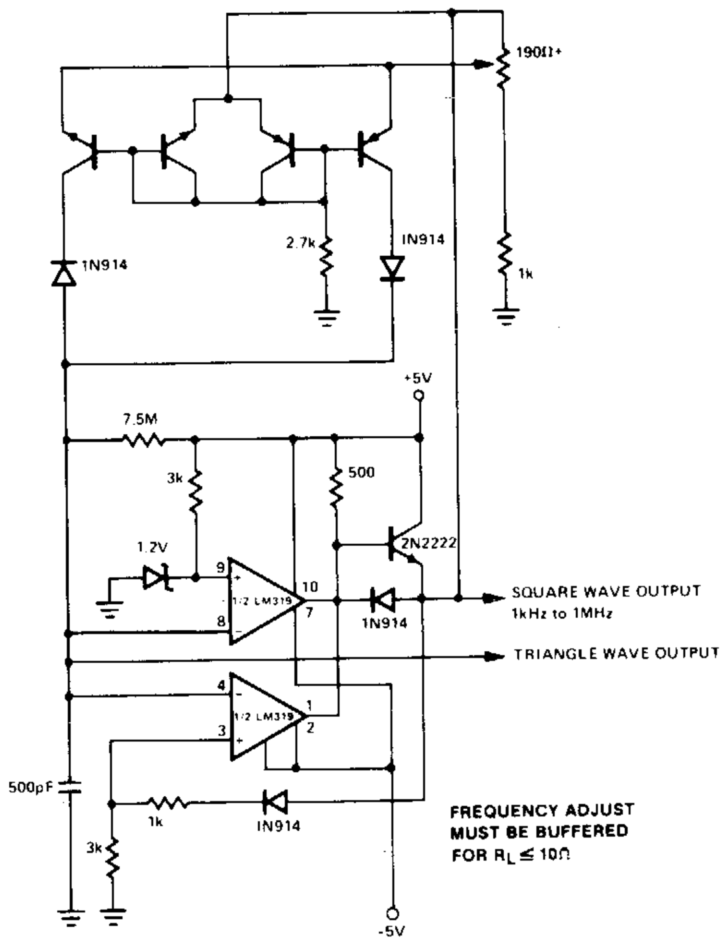
HANDS-ON ELECTRONICS

Fig. 66-12

Circuit Notes

The inexpensive 7404 hex-inverter has enough amplification to handle a wide range of transducers. Closing the key completes the battery circuit and applies four to five volts to the 7404. Bias for the first two inverter amps (U1a and U1b) comes from the two resistors, R1 and R2, connected between their inputs and outputs. The capacitor and rheostat (R3/C1) close the feedback loop from the input to the properly-phased output. The signal leaving U1b drives the remaining four inverter amplifiers, U1c through U1f; they, in turn, drive the phones or speakers. The volume control potentiometers, R4-R7, may have any value from 1500 ohms to 10,000 ohms. The smaller values work best when speakers, or low impedance phones, are used.

WIDE RANGE VARIABLE OSCILLATOR



SIGNETICS

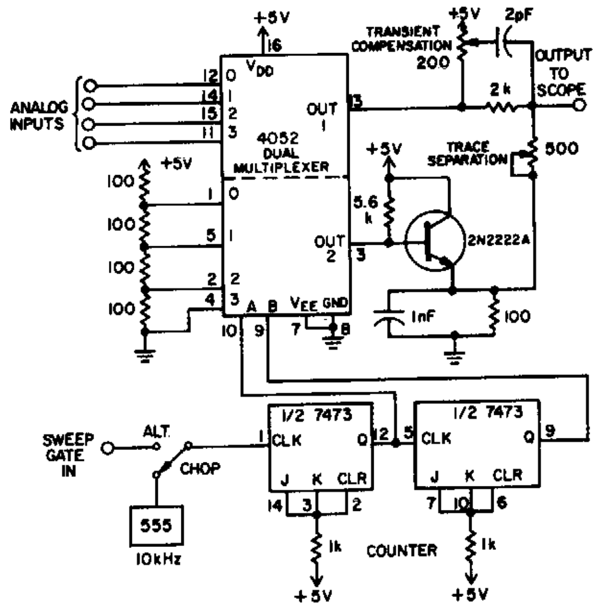
Fig. 66-13

67

Oscilloscope Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ANALOG MULTIPLEXER CONVERTS SINGLE-TRACE SCOPE TO FOUR-TRACE



ELECTRONIC DESIGN

Fig. 67-1

Circuit Notes

This adapter circuit, based on a dual four-channel analog multiplexer handles digital signals to at least 1 MHz, and analog signals at least through the audio range. The dual multiplexer's upper half selects one input for display. The lower half generates a staircase to offset the baselines of each channel, keeping them separate on the screen. The emitter-follower buffers the staircase, which is then summed with the selected signal. A two-bit binary counter addresses the CMOS 4052 multiplexer.

FET DUAL-TRACE SCOPE SWITCH

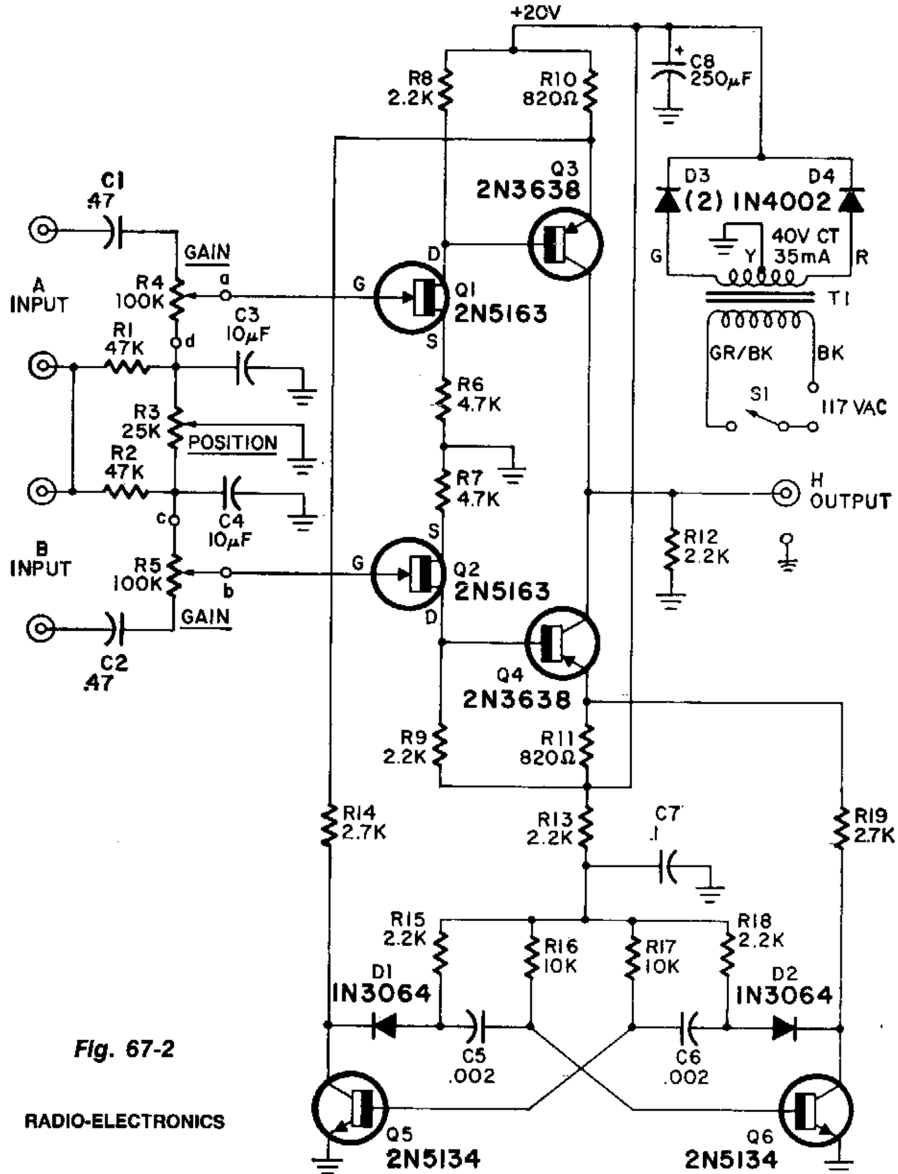


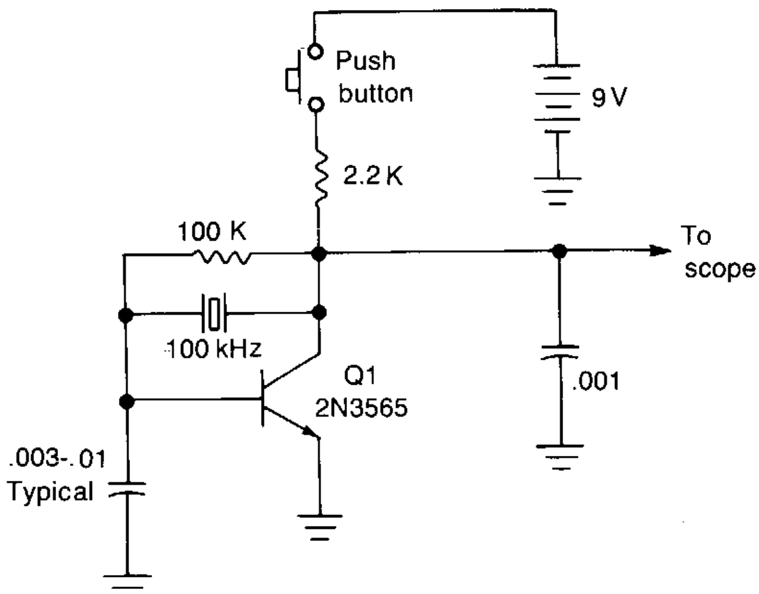
Fig. 67-2

RADIO-ELECTRONICS

Circuit Notes

The switcher output goes to the single vertical input of the scope, and a sync line from one of the inputs is taken to the scope's external-sync input. Frequency response of the input amplifiers is 300 kHz over the range of the gain controls. With the gain controls wide open so no attenuation of the signal takes place, the frequency response is up to 1 MHz.

SCOPE CALIBRATOR



WILLIAM SHEETS

Fig. 67-3

Circuit Notes

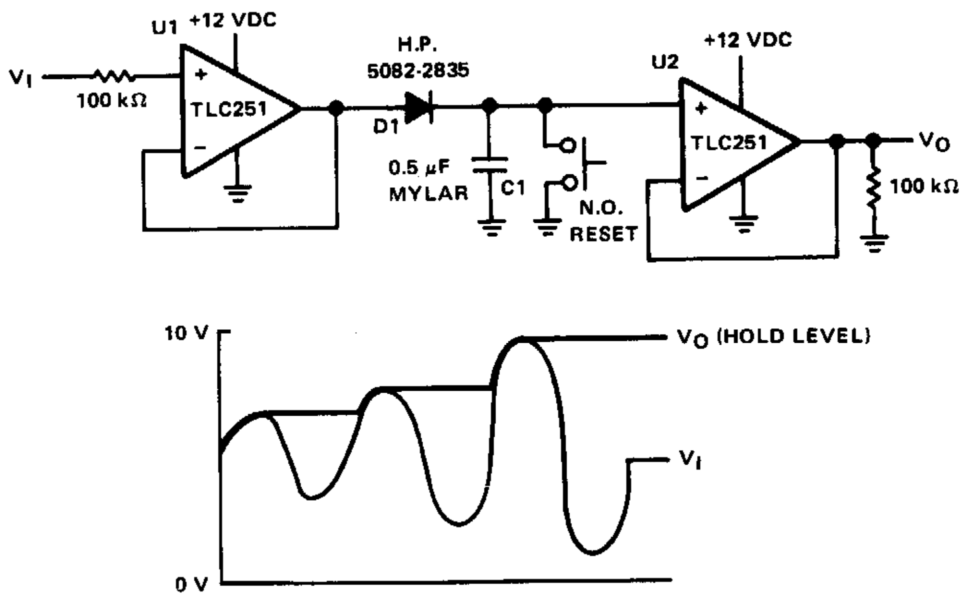
The calibrator operates on exactly 100 kHz providing a reference for calibrating the variable time base oscillator of general purpose scopes. For example, if the scope is set so that one cycle of the signal fills exactly 10 graticule divisions then each division represents 1 MHz, or 1 microsecond. If the scope is adjusted for 10 cycles on 10 graticule divisions. (1 cycle per division) then each division represents 100 kHz or 10 microseconds.

68

Peak Detector Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

POSITIVE PEAK DETECTOR



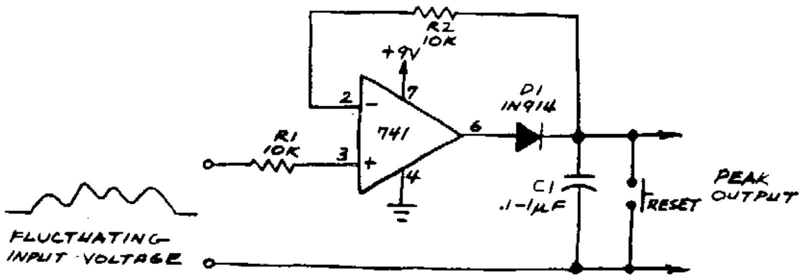
TEXAS INSTRUMENTS

Fig. 68-1

Circuit Notes

The purpose of the circuit is to hold the peak of the input voltage on capacitor C1, and read the value, V_O, at the output of U2. Op amps U1 and U2 are connected as voltage followers. When a signal is applied to V_I, C1 will charge to this same voltage through diode D1. This positive peak voltage on C1 will maintain V_O at this level until the capacitor is reset (shorted). Of course, higher positive peaks will raise this level while lower power peaks will be ignored. C1 can be reset manually with a switch, or electronically with an FET that is normally off. The capacitor specified for C1 should have low leakage and low dielectric absorption. Diode D1 should also have low leakage. Peak values of negative polarity signals may be detected by reversing D1.

PEAK DETECTOR



POPULAR ELECTRONICS

Fig. 68-2

Circuit Notes

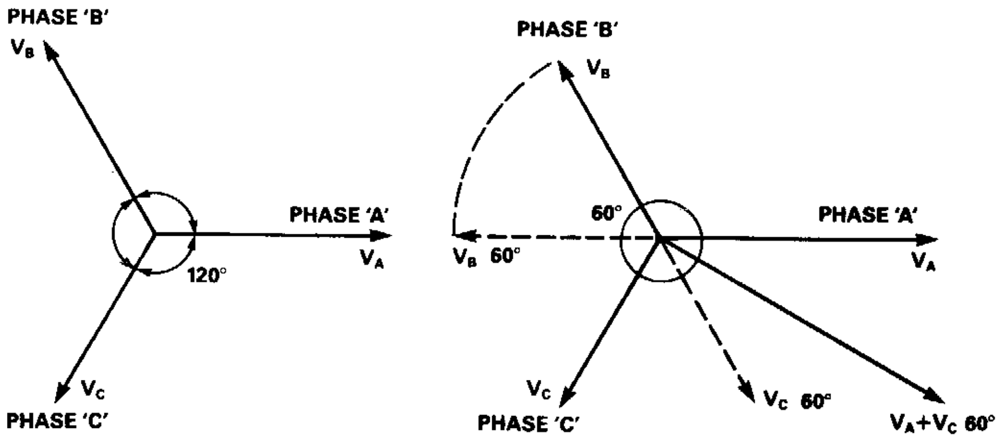
The comparator will charge C1 until the voltage across the capacitor equals the input voltage. If subsequent input voltage exceeds that stored in C1, the comparator voltage will go high and charge C1 to new higher peak voltage.

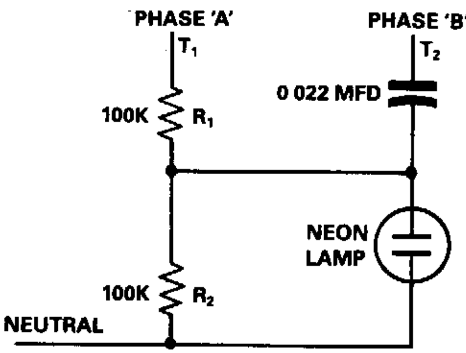
69

Phase Sequence Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RC CIRCUIT DETECTS PHASE SEQUENCE REVERSAL





TABLE

PHASE SEQUENCE	NEON INDICATOR	MOTOR MOTION
$V_A V_B V_C$	OFF	FORWARD
$V_A V_C V_B$	ON	REVERSE
$V_B V_A V_C$	ON	REVERSE
$V_B V_C V_A$	OFF	FORWARD
$V_C V_A V_B$	OFF	FORWARD
$V_C V_B V_A$	ON	REVERSE

ELECTRONIC ENGINEERING

Fig. 69-1

Circuit Notes

Assume the correct phase sequence to be V_A-V_B-V_C. The circuit terminals are connected such that T1 gets connected to phase A and T2 to phase B. The capacitor advances the voltage developed across R2 due to phase "B" by ~ 60°, while the voltages developed across it by phase "A" is in phase with V_A as shown in Fig. 69-1. The net voltage developed across R2 ~ zero, the neon lamp is not energized, thereby signaling correct phase sequence. If terminal T2 gets connected to phase C, a large voltage, K(V_A + V_C 60°), gets developed across R2, energizing the neon indicator to signal reverse phase sequence.

The motor terminals can be connected to the three phases in six different combinations. A three-phase motor will run in the forward direction for three such combinations, while for the other three it will operate in the reverse direction. As shown in the table, the circuit detects all three reverse combinations. This circuit can be wired into any existing motor starter where the operator can see whether the phase sequence has been altered, before starting the machine.

PHASE INDICATOR

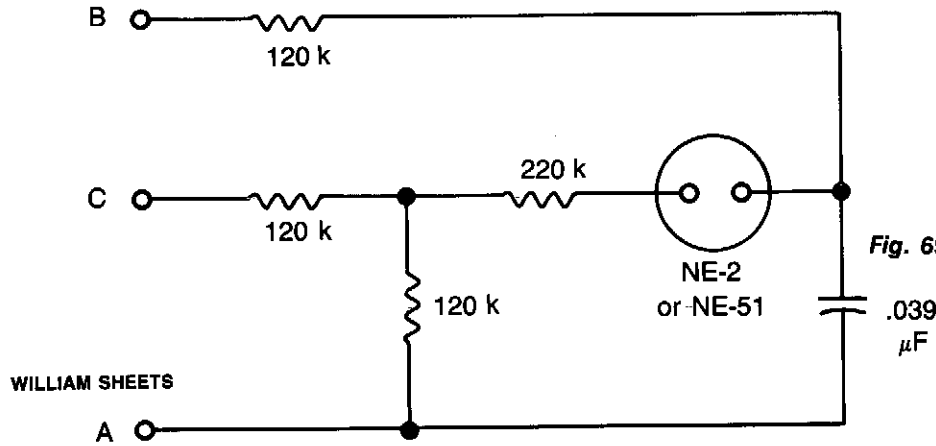


Fig. 69-2

Circuit Notes

The circuit provides a simple means of determining the succession of phases of a 3-phase 120 V source used in synchro work. Terminals A, B, and C are connected to the three terminals of the source to be checked. If the neon bulb lights, interchange any two leads; the light then extinguishes and A, B, and C indicate the correct sequence. If power on any one line is lost, the neon bulb will light. This feature may be useful for monitoring purposes.

PHASE SEQUENCE DETECTOR

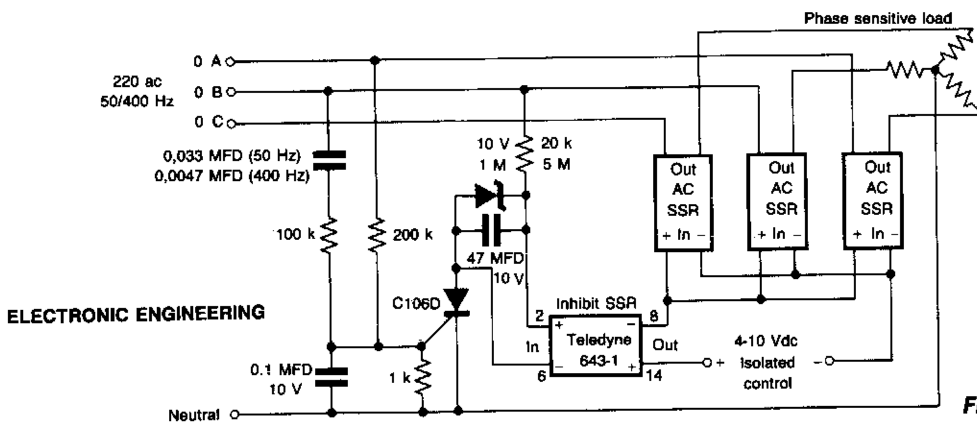
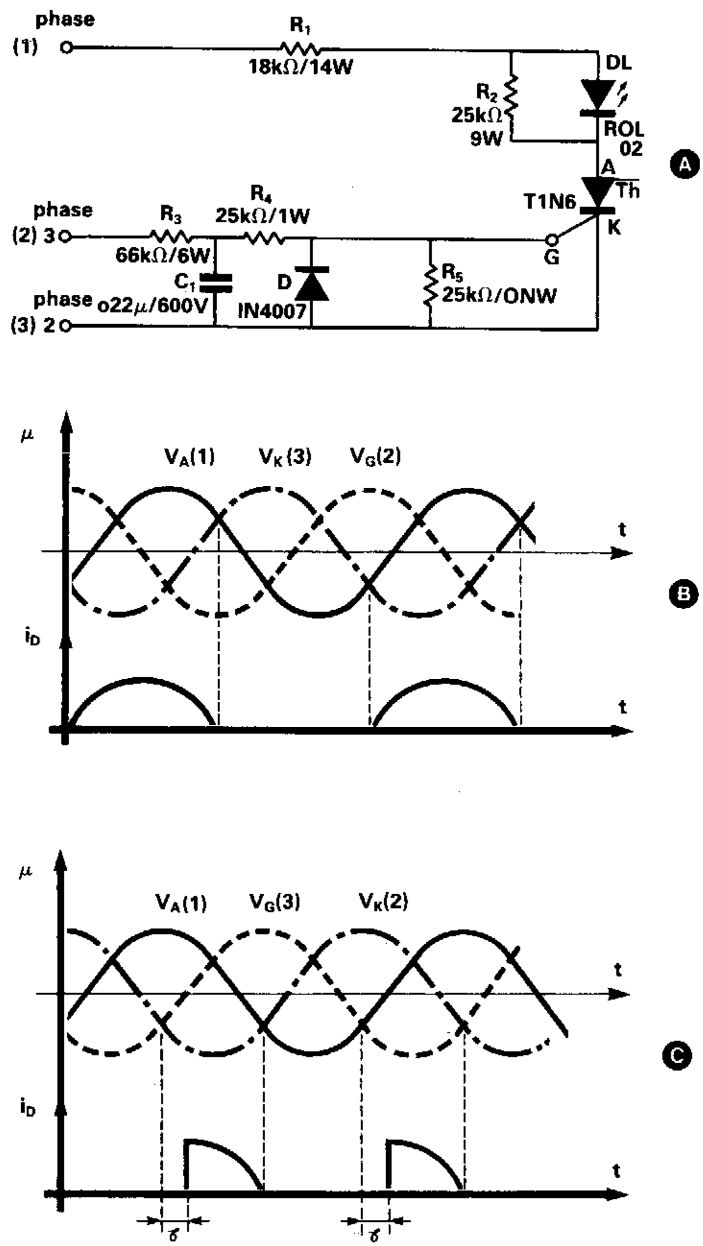


Fig. 69-3

Circuit Notes

This circuit prevents damage to the load due to incorrect phasing. The three power SSR's are only permitted to turn-on for a phase sequence of phase A leading phase B. If phase A lags phase B the input currents will cancel, causing the SCR and the inhibit SSR to remain off until the sequence is reversed. The inhibit SSR is included to maintain isolation at the input.

THREE PHASE TESTER



ELECTRONIC ENGINEERING

Fig. 69-4

THREE PHASE TESTER, Continued.

Circuit Notes

This simple three-phase tester, uses only a small current thyristor as a main element for testing the right or wrong succession of the three phases, and there is no need for a supplementary power supply.

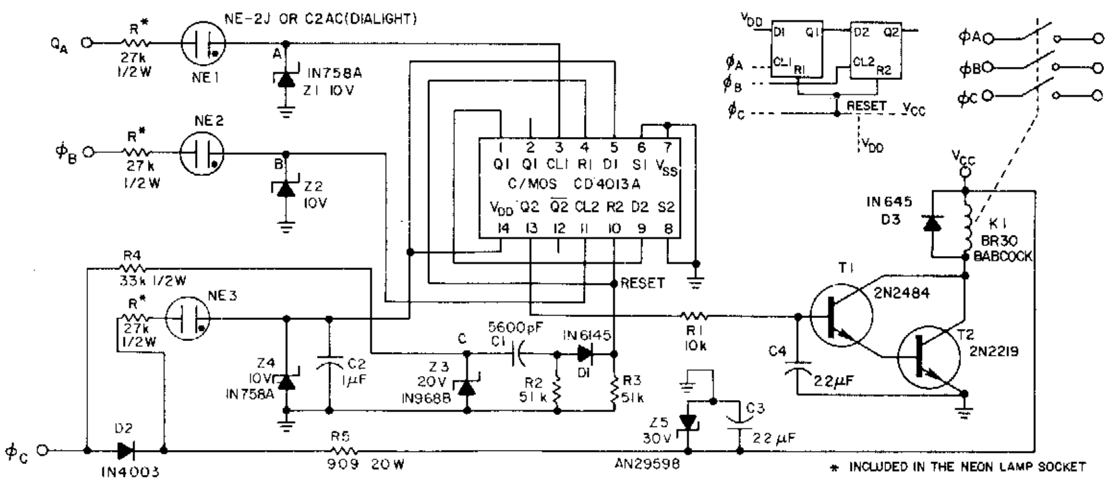
The basic circuit is shown in Fig. 69-4A. When connecting to the thyristor anode, grid, and cathode the three phases of the supply network in the sequence phase 1, phase 3, phase 2, are considered as correct, the mean value of the current through the thyristor is relatively high (since it is turned on during an entire half-period of one phase). The result is that the LED will emit a normal light.

The wave shapes for the three voltages and the current through the LED for this situation are shown in Fig. 69-4B.

If the three phases are not correctly connected—phase 1 to the anode, phase 2 to the grid, and phase 3 to the cathode, for instance—the thyristor will be turned on for a very short time and the LED will produce a very poor light. The wave shapes for this case are shown in Fig. 69-4C. The delay time is given by the R3-R1-R4 group.

When any of the three phases is missing, there is no current through the thyristor and the LED will emit no light.

PHASE-SEQUENCE DETECTOR II



ELECTRONIC DESIGN

Fig. 69-5

Circuit Notes

This circuit derives its supply voltage, V_{cc} and V_{dd} from ϕ_c. This factor, together with the neon lamps and zener diodes in the phase inputs, establishes 50% threshold that detects low voltage or absence of one or more phases. Relay K1 energizes for correct phase volts.

SIMPLE PHASE DETECTOR CIRCUIT

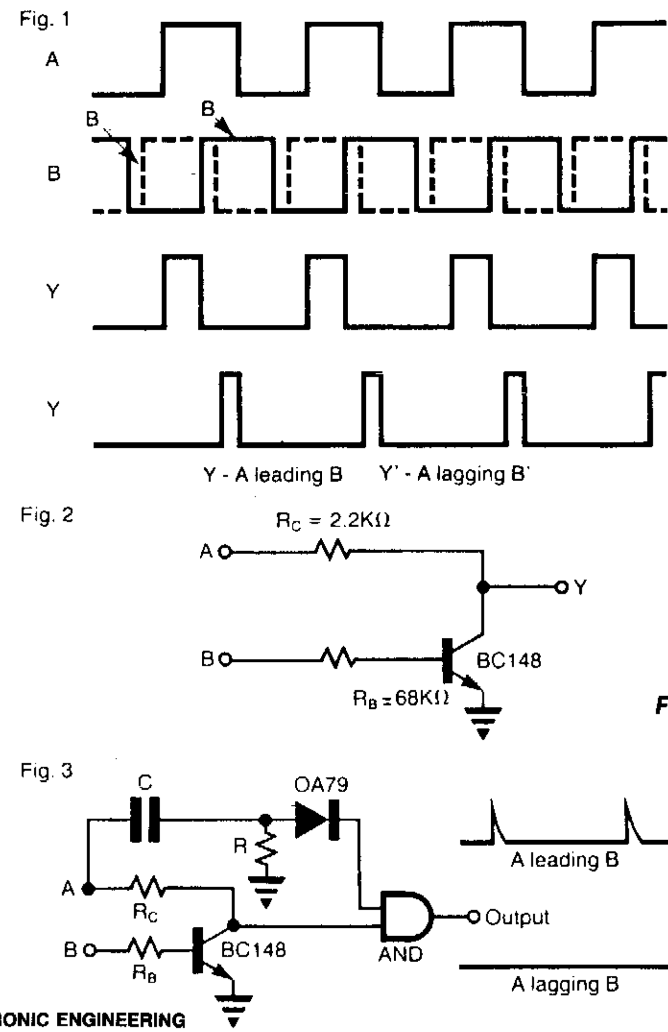


Fig. 69-6

ELECTRONIC ENGINEERING

Circuit Notes

The operation of the circuit is like an enabled inverter, that is, the output Y=B provided A is high. If A is low, output is low (independent of the state of B). When the signals A and B or B1 are connected to the inputs A and B of this gate the output Y is a pulse train signal (shown a Y or Y1) which has a pulse duration equal to the phase difference between the two signals. The circuit is directly suitable for phase difference measurement from zero to 180°. This performance is similar to the circuits like the Exclusive OR gate used for this purpose. With this method leading and lagging positions of the signals can also be found using an AND gate. Phase difference measured along with the leading and lagging information gives complete information about the phases of the two signals between zero and 360°.

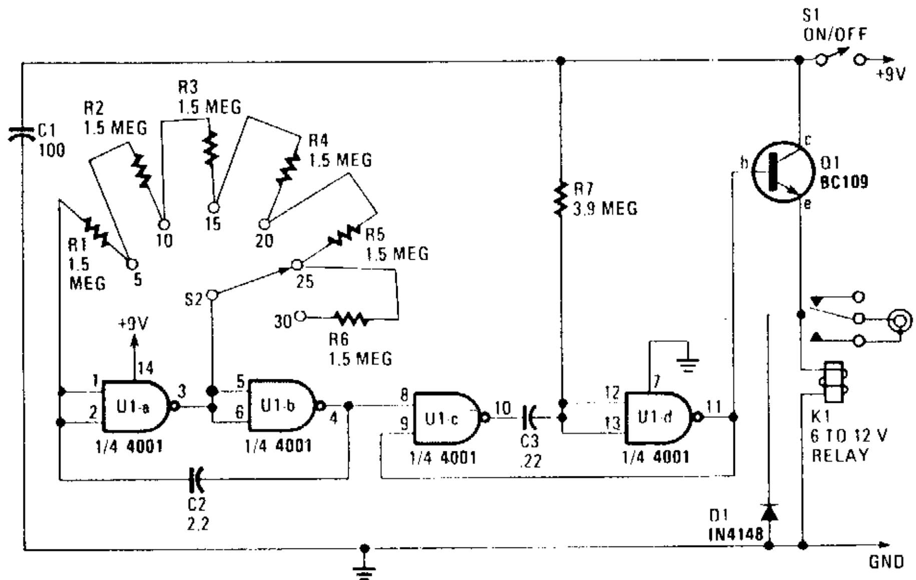
70

Photography-Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Auto-Advance Projector
Shutter-Speed Tester
Enlarger Timer
Contrast Meter
Electronic Flash Trigger
Sound Trigger for Flash Unit

AUTO-ADVANCE PROJECTOR



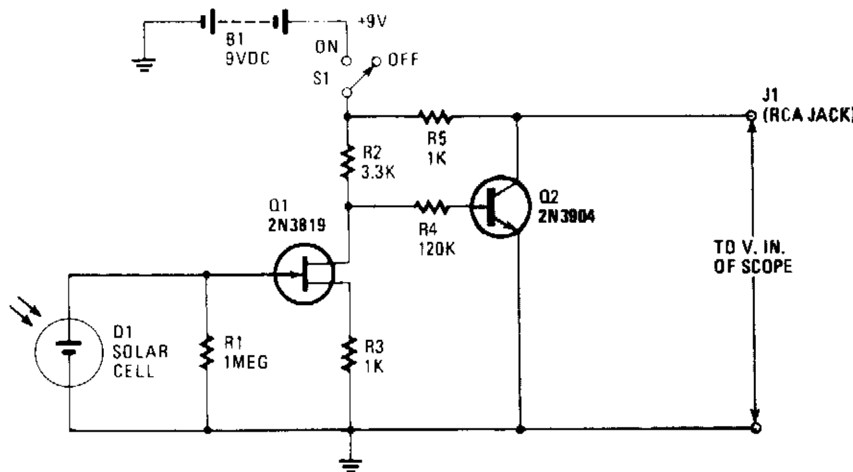
HANDS-ON ELECTRONICS

Fig. 70-1

Circuit Notes

The circuit is built around a 4001 quad two-input NOR gate, it provides switch selectable auto-advance times of 5, 10, 15, 20, 25 or 30 seconds through the remote-control socket of your projector. U1a and U1b form an astable multivibrator, with its operating frequency dependent on the number of timing resistors switched into the circuit via S2. The frequency is about one cycle for every five seconds with a single timing resistor, one every ten seconds with two resistors, etc., providing six switched time intervals. The output of the astable at pin 4 of U1b is fed to the input of a monostable multivibrator, consisting of the second pair of gates, U1c and U1d. R7 and C3 are the timing components; they set the length of the (positive) output pulse of the monostable at a little more than half a second. The monostable is triggered by each positive-going input it receives from the astable. The output from the monostable therefore, consists of a series of short pulses, the interval between the pulses being controlled using S2. The output of the monostable (at pin 11) controls a relay by way of Q1, which is configured as an emitter-follower buffer stage. The projector is controlled via the normally-open contacts of relay K1. When the output of the monostable goes positive, the relay contacts close, triggering the slide-change mechanism of the projector. The monostable assures that the power to the relay is applied only briefly by the timer, so that multiple operation of the projector is avoided.

SHUTTER-SPEED TESTER



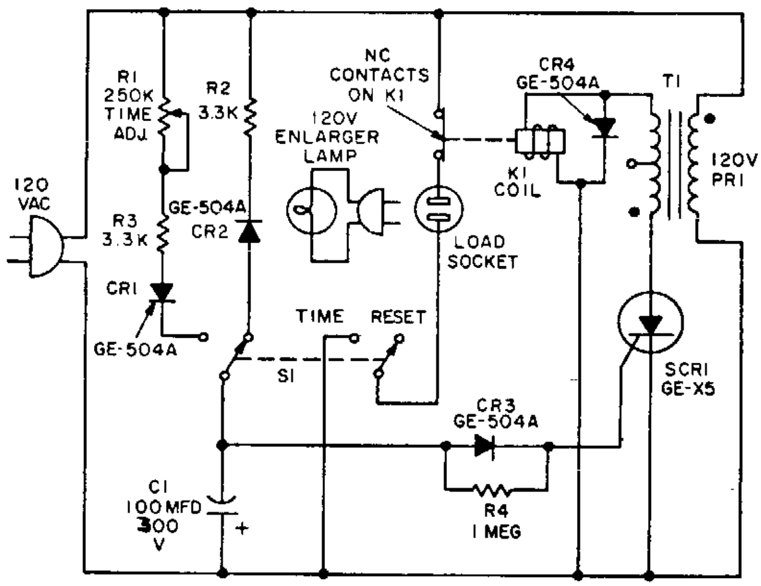
HANDS-ON ELECTRONICS

Fig. 70-2

Circuit Notes

The solar cell is connected across the input of the FET (field-effect transistor), Q1, so that it will produce positive dc voltage to the gate when activated by light shining through the open shutter, decreasing the negative gate-source bias already established by the source resistor, and causes an increase in drain current. The drain voltage goes more negative which causes a decrease in Q2's base current. Q2's collector current decreases, and its collector voltage becomes more positive. There is an amplified positive-going voltage output at the collector, and it's applied directly to the oscilloscope's vertical input, producing a waveform that is displaced vertically whenever light strikes the cell.

ENLARGER TIMER



Parts List

C1 — 100-mfd, 300-volt electrolytic capacitor
CR1 thru CR4 — GE-504A rectifier diode
K1 —12 -volt a-c relay (Potter & Brumfield No. MR5A, or equivalent)
R1 — 250K-ohm, 2-watt potentiometer
R2, R3 — 3.3K-ohm, 1/2-watt resistor
R4 — 1-megohm, 1/2-watt resistor
S1 — DPDT toggle switch
SCR1 — GE-X5 silicon controlled rectifier
T1 — Filament transformer: primary, 120-volts a-c; secondary, 12.6-volts center tapped (Triad F25X, or equivalent)
Line cord, vectorboard, minibox etc.

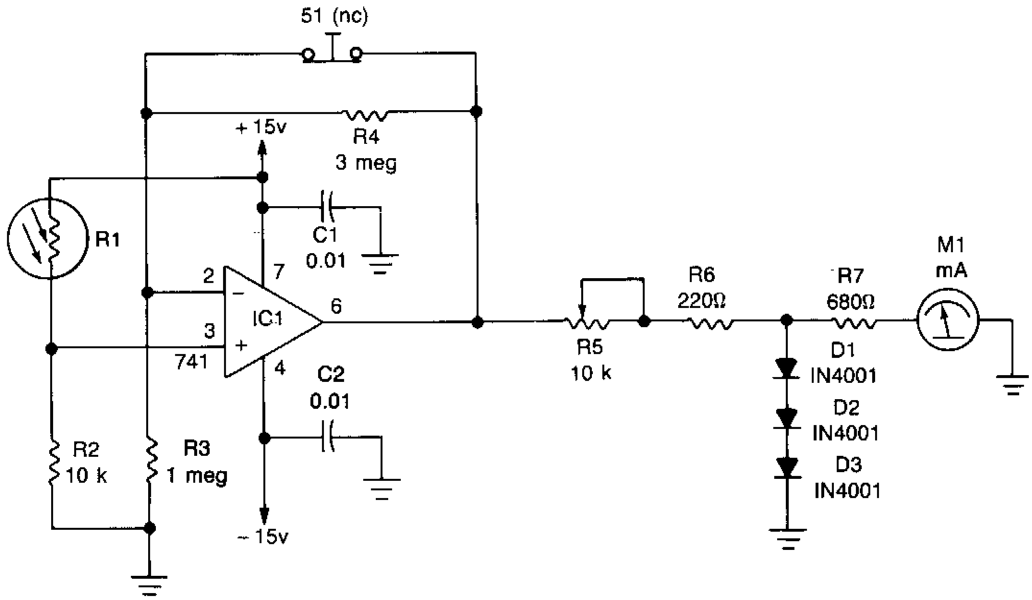
GENERAL ELECTRIC

Fig. 70-3

Circuit Notes

This precision, solid state, time delay circuit has delayed *off* and delayed *on* switching functions that are interchangeably available by simply interchanging the relay contacts.

CONTRAST METER



RADIO ELECTRONICS

Fig. 70-4

Circuit Notes

One leg of the photocell (R1) is tied to the +15 volt supply and the other end is connected to ground through resistor R2, forming a voltage-divider network. The non-inverting input of the 741 op amp, IC1, is tied to the junction formed by R1 and R2, while its inverting input is grounded through resistor R3. When switch S1 is pressed, another divider network is formed, reducing the voltage applied to the inverting input of the op amp. When light hits the photocell its resistance begins to decrease causing a greater voltage drop across R2 and a higher voltage to be presented to the non-inverting input of IC1. This causes IC1 to output a voltage proportional to the two inputs. The circuit gives a meter reading that depends on the intensity of light hitting photocell R1; therefore, R1 should be mounted in a bottle cap so that the light must pass through a 3/16 inch hole. Potentiometer R5 is used to adjust the circuit for the negative you're working with.

ELECTRONIC FLASH TRIGGER

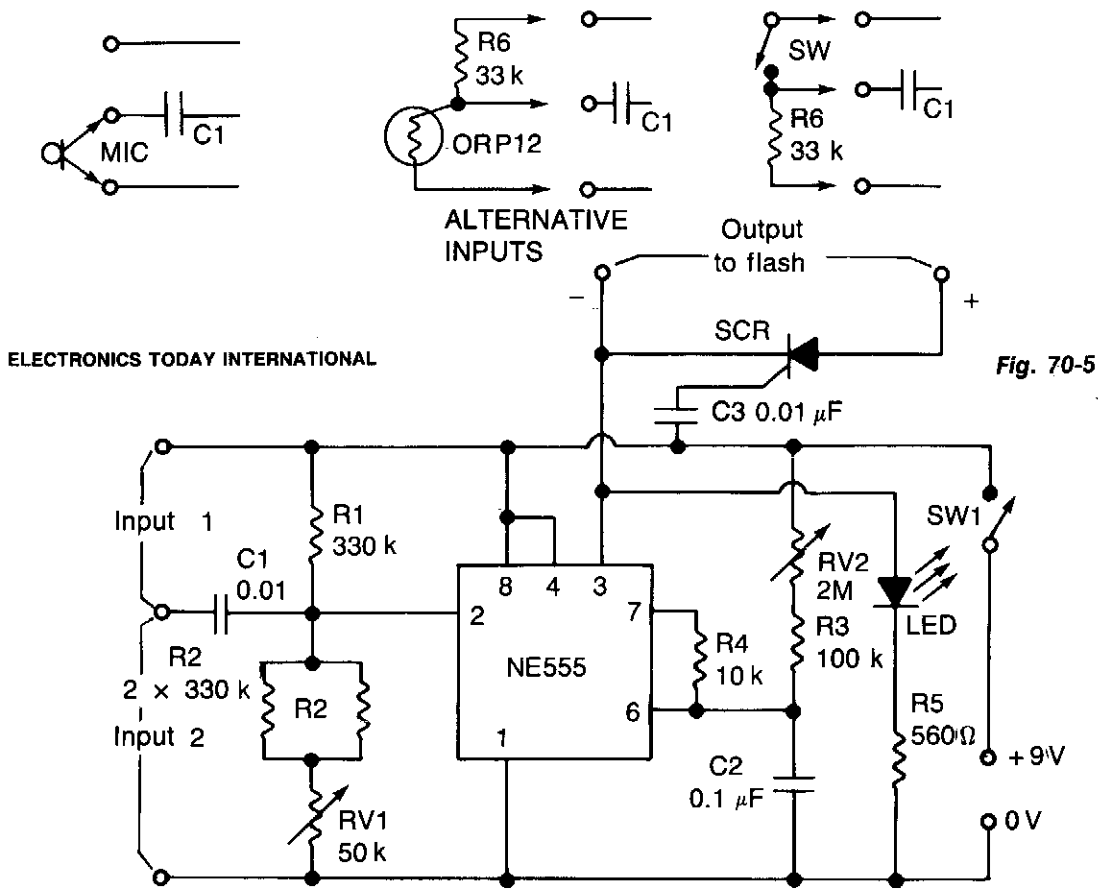
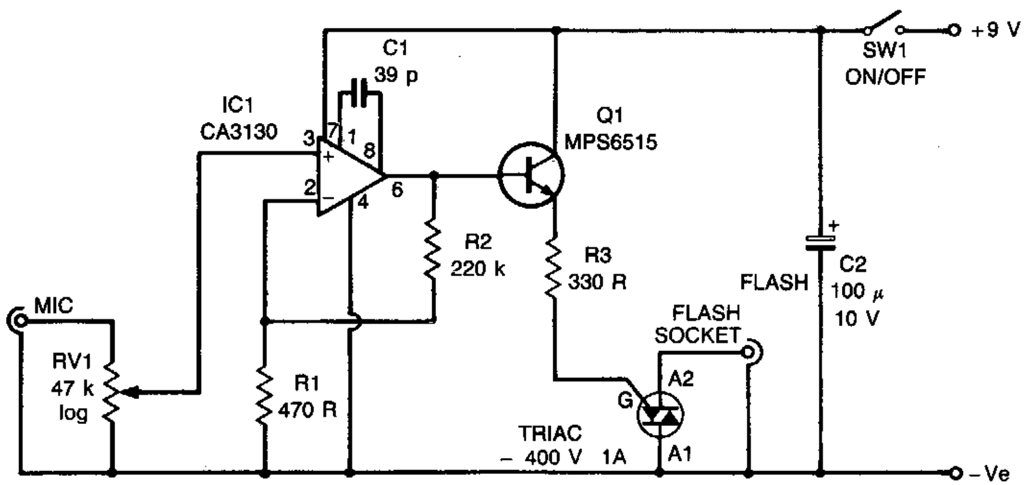


Fig. 70-5

Circuit Notes

A negative pulse at the input is fed via capacitor C1 to the input pin (2) of the IC. Pin 2 is held slightly above its triggering voltage of 1/3 V_{cc} by the voltage divider comprising R1, R2 and RV1. The negative pulse triggers the IC and the output (pin 3) goes high for a time period controlled by RV2, R3 and C2. When the output goes low again at the end of the time interval, capacitor C3 charges through the gate cathode circuit of the SCR switching it on and firing the flash. Capacitor C1 isolates the input from the voltage divider so that the unit isn't sensitive to the dc level at the input. RV1 acts as a sensitivity control by allowing the voltage to be adjusted to a suitable level so that the input signal will trigger the IC. Resistor R4 limits the discharge current from C2 at the end of the timing cycle protecting the IC. The LED and its protective resistor R5 act as an indicator to show that the unit has triggered, simplifying the setting up process and minimizing the number of times the strobe has to be fired. This means that the strobe needn't be fired until a photo is to be taken.

SOUND TRIGGER FOR FLASH UNIT



ELECTRONICS TODAY INTERNATIONAL

Fig. 70-6

Circuit Notes

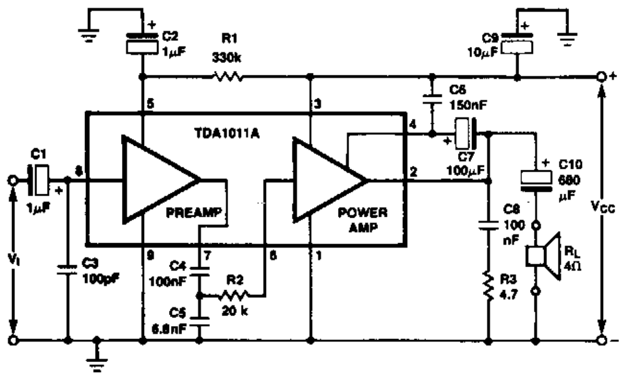
The circuit is based on operational amplifier IC1 used in the noninverting amplifier mode. R1 and 2 set the gain at about 500. RV1 (sensitivity) biases the noninverting input to the negative supply. Q1 provides the relatively high trigger current required by the triac. When a signal is received by the microphone, the signals are amplified (by IC1). The triac is triggered and a low resistance appears across its A1 and A2 terminals which are connected via the flashlead to the strobe. The circuit operates almost instantly, giving very little delay between the commencement of the sound and the flashgun being triggered.

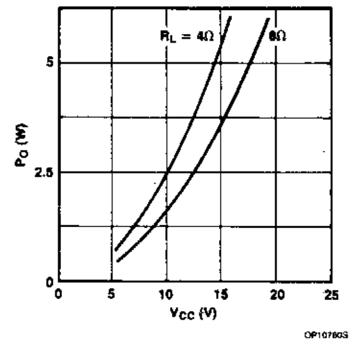
71

Power Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

2 TO 6 W AUDIO AMPLIFIER WITH PREAMPLIFIER





NOTES:
d_{TOT} = 10%; typical values. The available output power is 5% higher when measured at Pin 2 (due to series resistance of C1).

Output Power Across R_L as a Function of Supply Voltage with Bootstrap

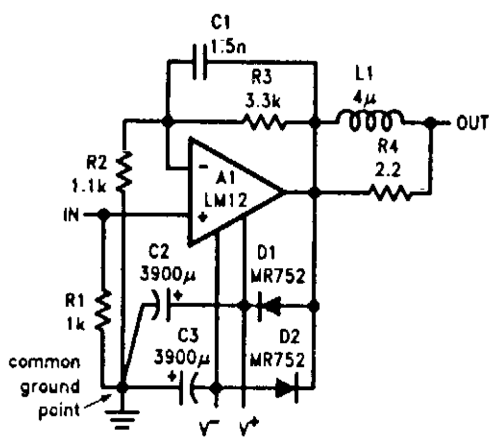
SIGNETICS

Fig. 71-1

Circuit Notes

The monolithic integrated audio amplifier circuit is especially designed for portable radio and recorder applications and delivers up to 4 W in a 4 ohm load impedance.

AUDIO POWER AMPLIFIER



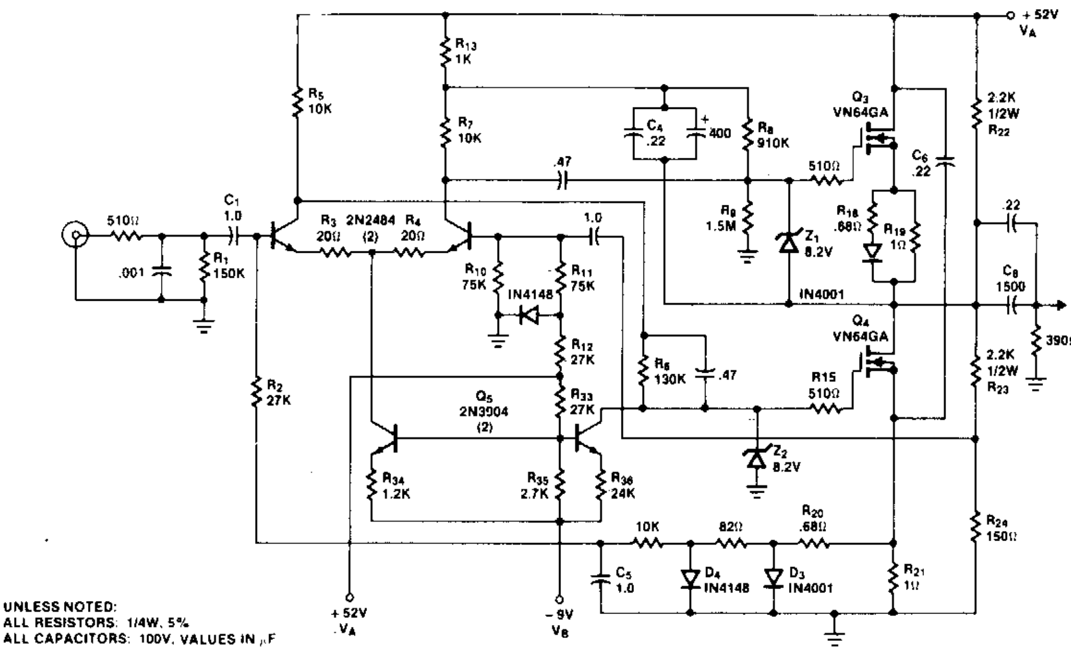
NATIONAL SEMICONDUCTOR CORP.

Fig. 71-2

Circuit Notes

Output-clamp diodes are mandatory because loudspeakers are inductive loads. Output LR isolation is also used because audio amplifiers are usually expected to handle up to 2 μF load capacitance. Large, supply-bypass capacitors located close to the IC are used so that the rectified load current in the supply leads does not get back into the amplifier, increasing high-frequency distortion. Single-point grounding for all internal leads plus the signal source and load is recommended to avoid ground loops that can increase distortion.

25 WATT AMPLIFIER



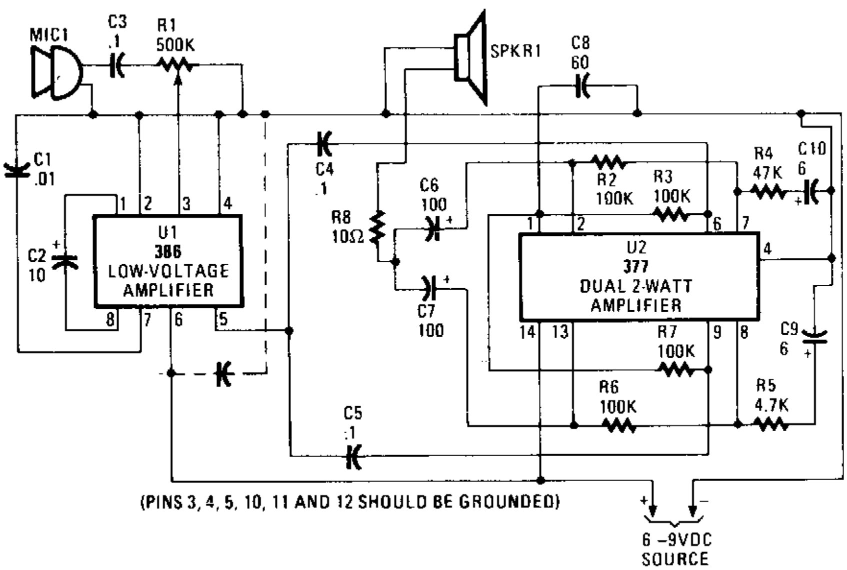
SILICONIX, INC.

Fig. 71-3

Circuit Notes

Transistors are used for current sources. Base drive for these transistors is derived from the main power supply V_A, so that their collector current is proportional to the rail voltage. This feature holds the voltage on the diff-amp collectors close to $V_A/2$. The sensitivity of I_Q to V_A is about 3.4 mA/volt when V_B is held constant; the sensitivity of I_Q to V_B is -15 mA/volt when V_A is held constant. In a practical amplifier with a non-regulated supply, variations in power output will cause fluctuations in V_A, but will not affect V_B; therefore, having I_Q increase slightly with power output will tend to compensate for the 3.4 mA/volt I_QV_A sensitivity. In the case of line voltage variations, since V_A is about five times V_B, the sensitivities tend to cancel, leaving a net sensitivity of about 2 mA/volt.

BULL HORN

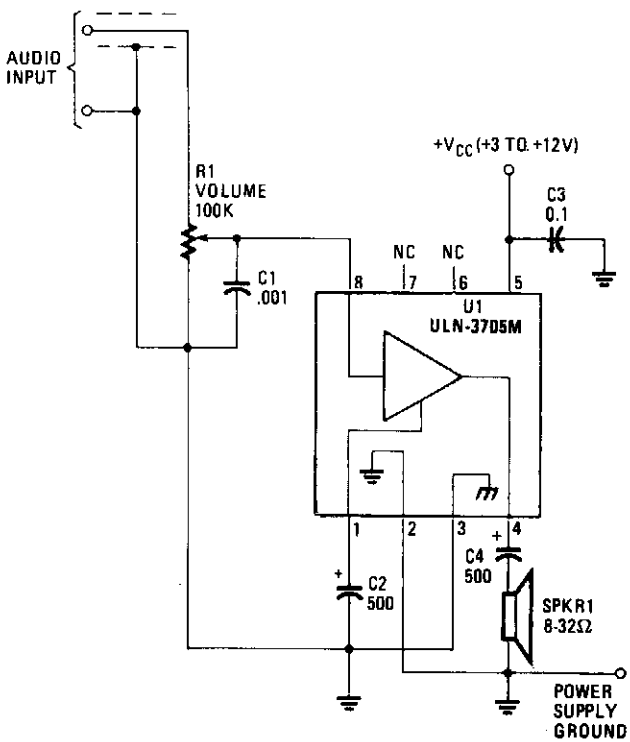


HANDS-ON ELECTRONICS *Fig. 71-4*

Circuit Notes

The input audio signal is fed to pin 3 of U1, an LM386 low-voltage amplifier, via C3 and R1. Potentiometer R1 sets the drive or volume level. U1, which serves as a driver stage, can be set for a gain of from 20 to 200. The output of U1 at pin 5 is fed to U2—a 377 dual two-watt amplifier connected in parallel to produce about four watts of output power—at pins 6 and 9 via C4 and C5. Frequency stability is determined by R2, R4, and C10 on one side, and the corresponding components R6, R5, and C9 on the other side. The outputs of the two amplifiers (at pins 2 and 13) are capacitively coupled to SPKR1 through C6 and C7.

LOW POWER AUDIO AMPLIFIER



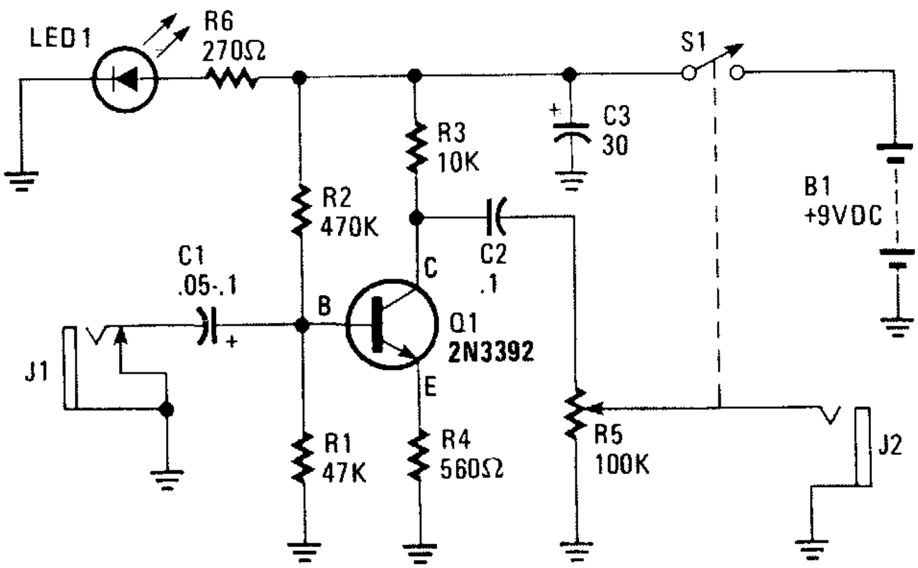
HANDS-ON ELECTRONICS

Fig. 71-5

Circuit Notes

The amplifier operates from supplies ranging up to 12 volts, and operates (with reduced volume) from supply voltages as low as 1.8 volts without having distortion rise to unacceptable levels. (Its power requirements make it suitable for solar-cell application.) Components external to the integrated circuit, U1, consist of four capacitors and a potentiometer for volume control. Capacitor C3 is for decoupling, low-frequency roll-off, and power-supply ripple rejection. Capacitor C4 is an electrolytic type that couples the audio output to an 8 to 32 ohm speaker that is efficient.

AUDIO BOOSTER



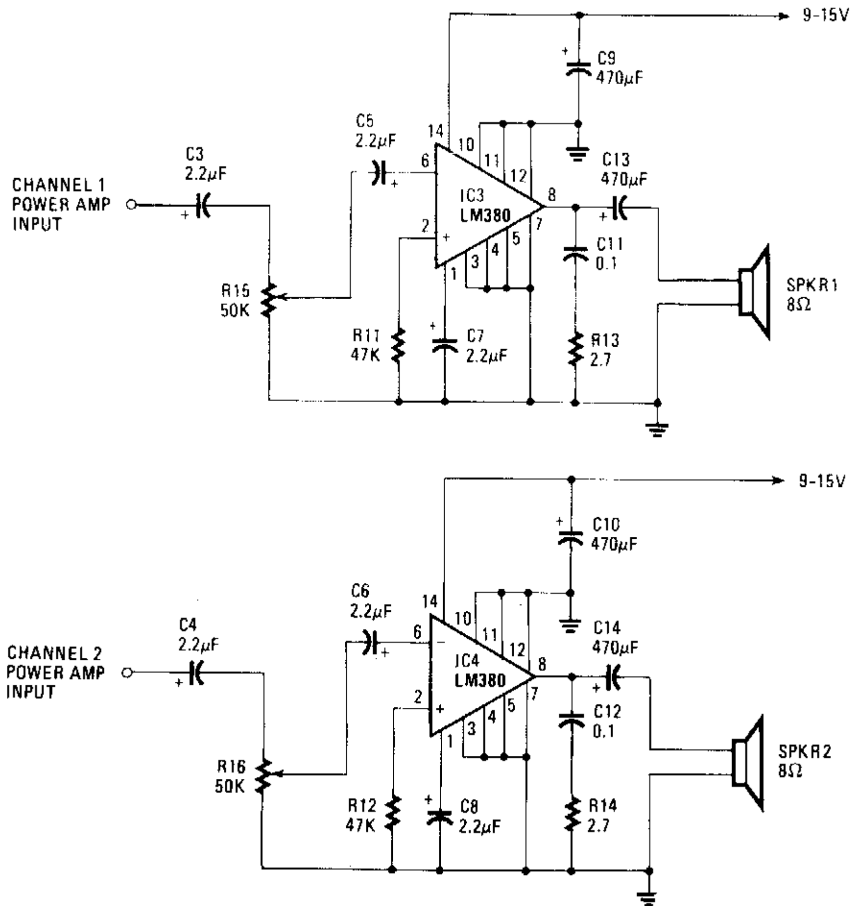
HANDS-ON ELECTRONICS

Fig. 71-6

Circuit Notes

The amplifier's gain is nominally 20 dB. Its frequency response is determined primarily by the value of just a few components—primarily C1 and R1. The values of the schematic diagram provide a response of ± 3.0 dB from about 120 Hz to better than 20,000 Hz. Actually, the frequency response is ruler flat from about 170 Hz to well over 20,000 Hz; it's the low end that deviates from a flat frequency response. The low end's roll-off is primarily a function of capacitor C1 (since R1's resistive value is fixed). If C1's value is changed to 0.1 μF, the low end's corner frequency—the frequency at which the low-end roll-off starts—is reduced to about 70 Hz. If you need an even deeper low-end roll-off, change C1 to a 1.0 μF capacitor; if it's an electrolytic type, make certain that it's installed into the circuit with the correct polarity, with the positive terminal connected to Q1's base terminal.

WALKMAN AMPLIFIER



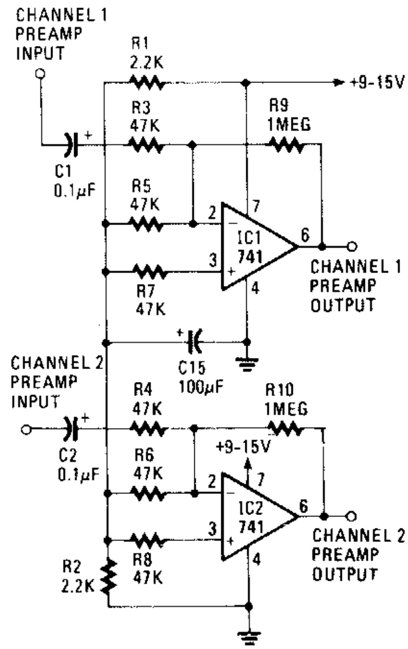
RADIO-ELECTRONICS

Fig. 71-7

Circuit Notes

The gain of the low-cost IC is internally fixed so that it is not less than 34 dB (50 times). A unique input stage allows input signals to be referenced to ground. The output is automatically self centering to one half the supply voltage. The output is also short-circuit proof with internal thermal limiting. With a maximum supply of 15 volts and an 8 ohm load, the output is around 1.5 watts per channel. The input stage is usable with signals from 50 mV to 500 mV rms. If the amplifier is to be used with a source other

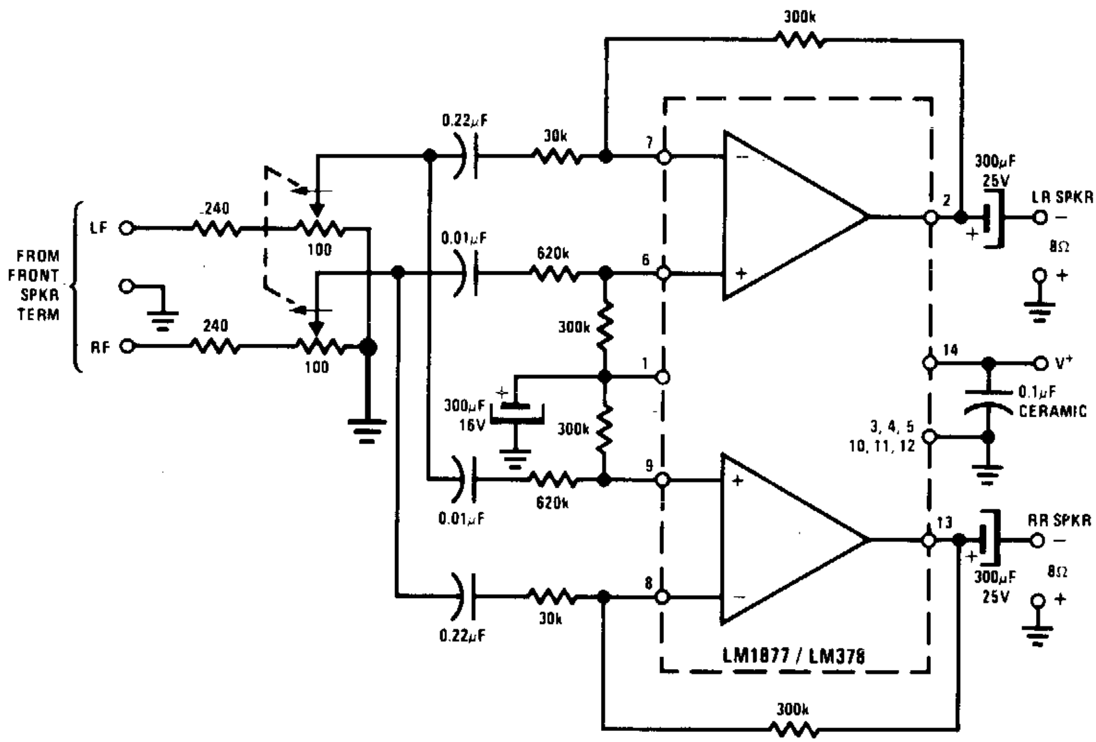
WALKMAN AMPLIFIER, Continued.



-THE PREAMP. **If you wish to amplify low-level signals, such as the output of a turntable, the signal will first have to be fed to the preamp shown here.**

than a personal stereo, such as a phonograph or an electric guitar, some type of preamplifier is required. A suitable circuit is shown. In that circuit, two 741 op amps have been configured as input amplifiers. Their input stages referenced to a common point—half the supply voltage. That voltage is derived from a voltage divider made up of R1 and R2, two 2.2 k resistors. The gain of each of the 741's has been fixed at 21 by the input resistors (R9, R10). Input capacitors, C1 and C2, are used to filter out any dc component from the input signal.

REAR SPEAKER AMBIENCE (4-CHANNEL) AMPLIFIER

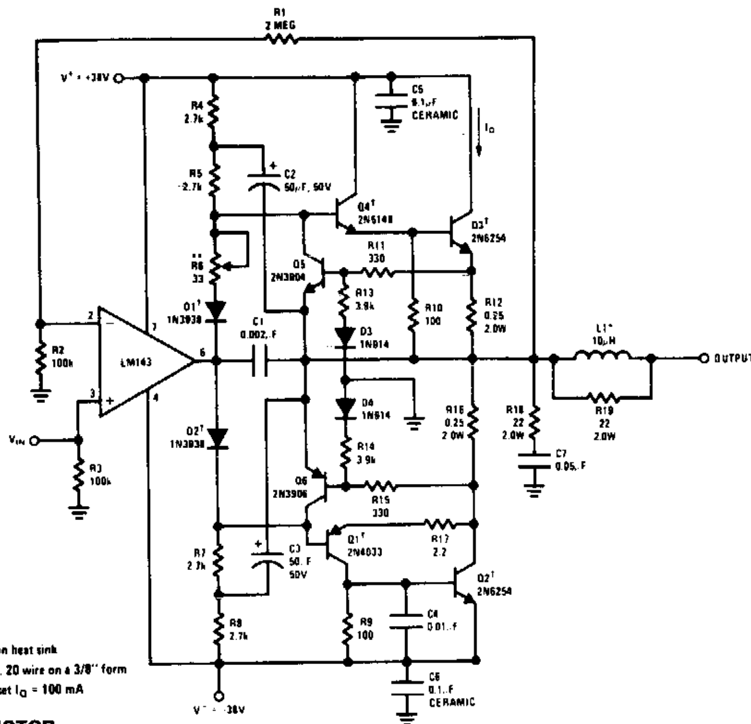


NATIONAL SEMICONDUCTOR CORP. *Fig. 71-8*

Circuit Notes

Rear channel "ambience" can be added to an existing stereo system to extract a difference signal (R - L or L - R) which, when combined with some direct signal (R or L), adds fullness, or "concert hall realism" to the reproduction of recorded music. Very little power is required at the rear channels, hence an LM1877 suffices for most "ambience" applications. The inputs are merely connected to the existing speaker output terminals of a stereo set, and two more speakers are connected to the ambience circuit outputs. The rear speakers should be connected in the opposite phase to those of the front speakers, as indicated by the +/− signs.

90 W AUDIO POWER AMPLIFIER WITH SAFE AREA PROTECTION



†Put on common heat sink

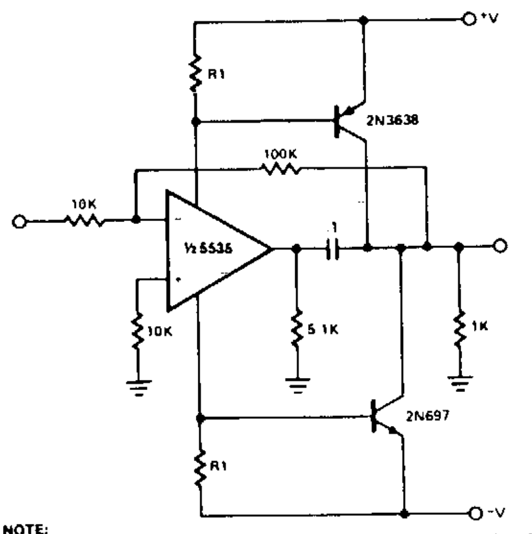
*34 turns of no. 20 wire on a 3/8" form

**Adjust R6 to set I_Q = 100 mA

NATIONAL SEMICONDUCTOR

Fig. 71-9

POWER AMPLIFIER



NOTE:
All resistor values are in ohms.

SIGNETICS

Fig. 71-10

Circuit Notes

For most applications, the available power from op amps is sufficient. There are times when more power handling capability is necessary. A simple power booster capable of driving moderate loads uses an NE5535 device. Other amplifiers may be substituted only if R1 values are changed because of the I_{CC} current required by the amplifier. R1 should be calculated from the expression

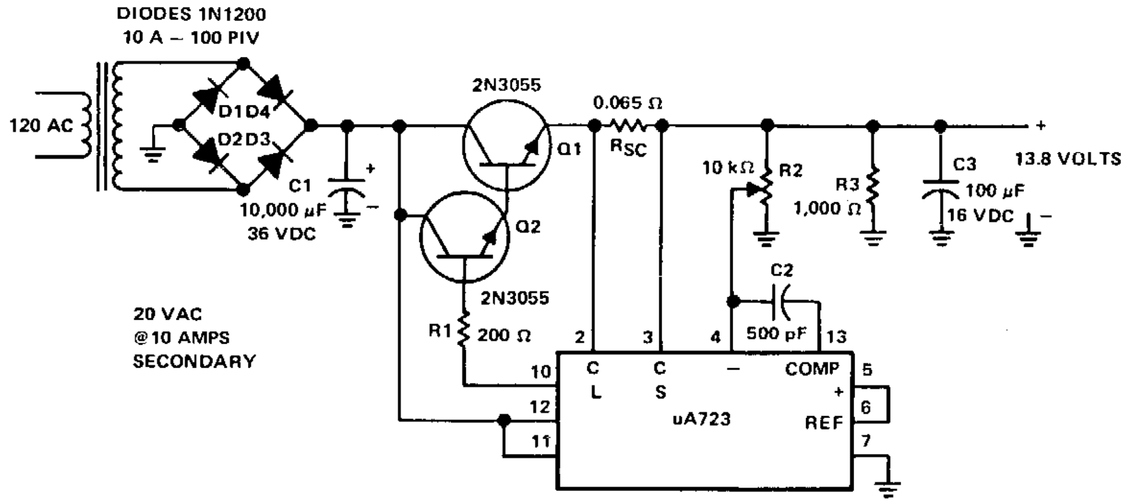
$$R1 = \frac{600\ mV}{I_{CC}}$$

72

Power Supply Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-AMP REGULATED POWER SUPPLY FOR OPERATING MOBILE EQUIPMENT



TEXAS INSTRUMENTS

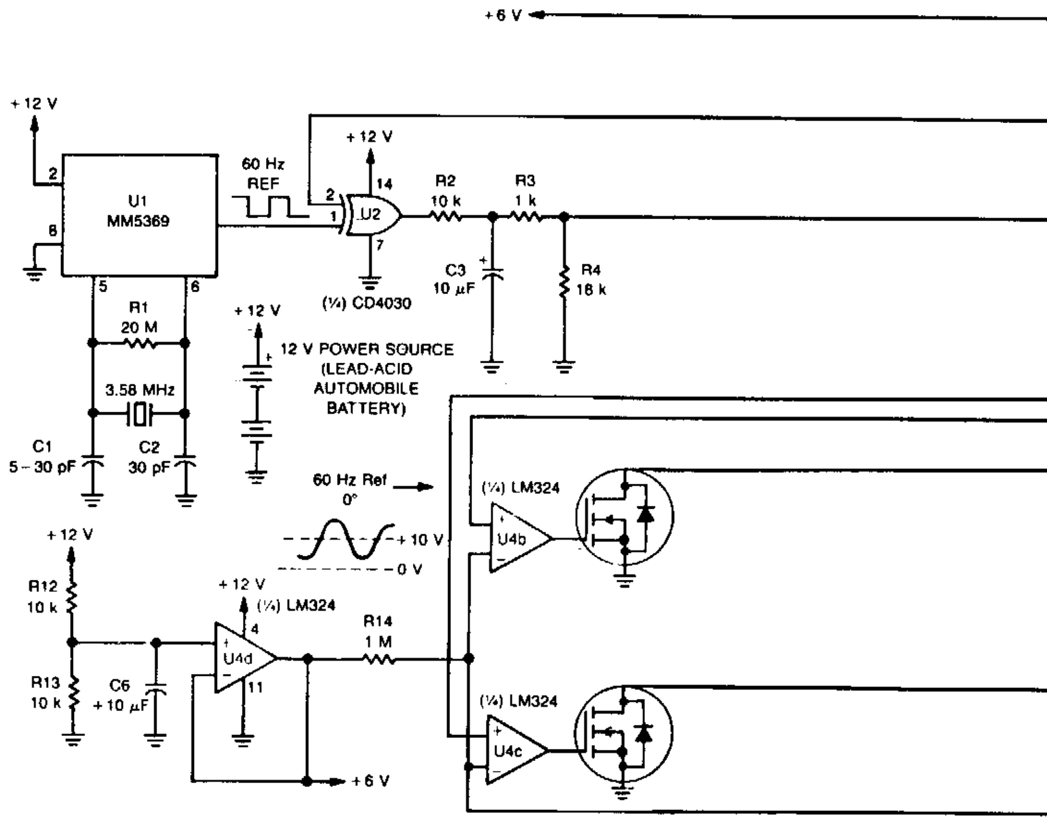
Fig. 72-1

Circuit Notes

This supply is powered by a transformer operating from 120 Vac on the primary and providing approximately 20 Vac on the primary, and providing approximately 20 Vac on the secondary. Four 10-A diodes with a 100 PIV rating are used in a full-wave bridge rectifier. A 10,000 μF/36 Vdc capacitor completes the filtering, providing 28 Vdc. The dc voltage is fed to the collectors of the Darlington connected 2N3055's. Base drive for the pass transistors is from pin 10 of the μA723 through a 200 ohm current limiting resistor, R1. The reference terminal (pin 6) is tied directly to the non-inverting input of the error amplifier (pin 5), providing 7.15 V for comparison.

The inverting input to the error amplifier (pin 4) is fed from the center arm of a 10 k ohm potentiometer connected across the output of the supply. This control is set for the desired output voltage of 13.8 V. Compensation of the error amplifier is accomplished with a 500 pF capacitor connected from pin 13 to pin 4. If the power supply should exceed 8 A or develop a short circuit, the μA723 regulator will bias the transistors to cutoff and the output voltage will drop to near zero until the short circuit condition is corrected.

UNINTERRUPTIBLE POWER SUPPLY FOR PERSONAL COMPUTERS



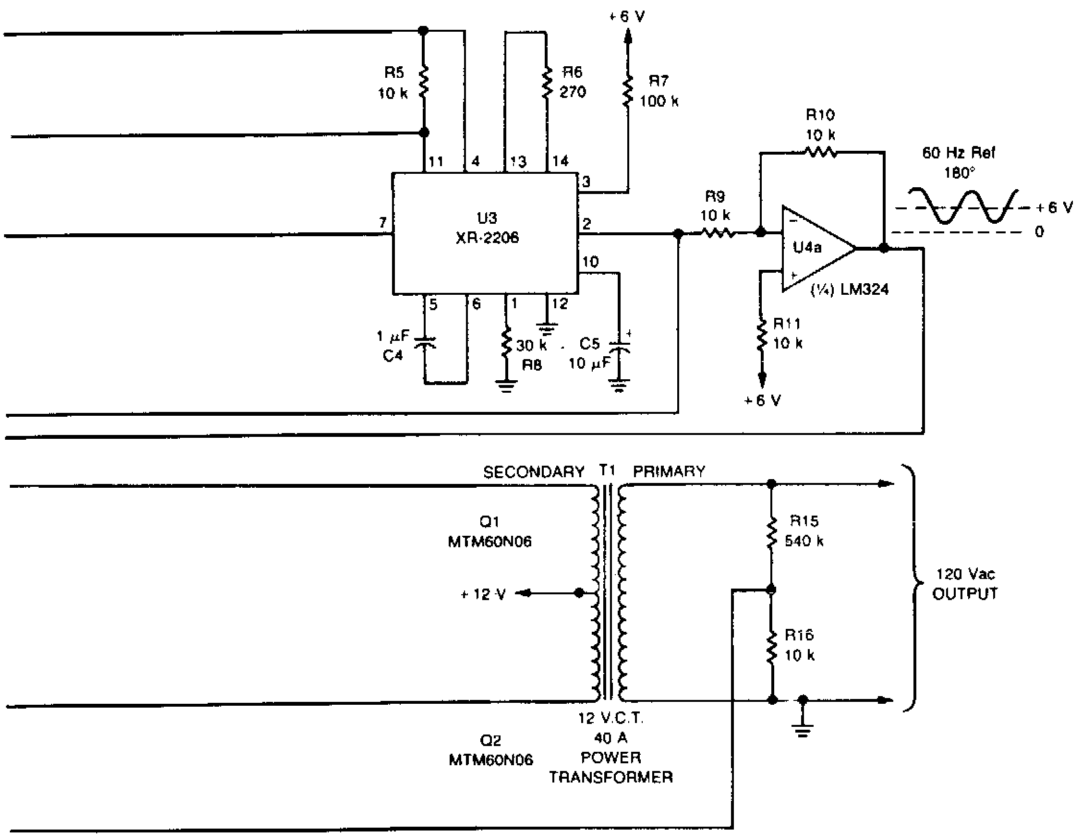
MOTOROLA

Fig. 72-2

Circuit Notes

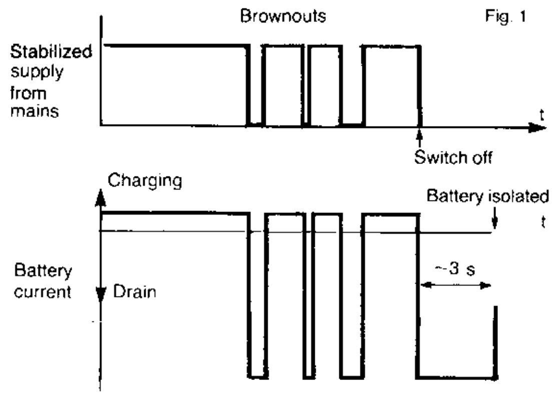
The UPS is basically an ac inverter that is powered by a 12-V, lead-acid automobile battery. During power outages, it can supply several minutes of power for an average personal computer. It incorporates a crystal-controlled 60 Hz time base, so that a computer with a real time clock can maintain its accuracy. It isolates the ac line from the computer, so it can be used to operate sensitive electronic equipment on noisy power sources.

Two MTM60N06 Power FETs (Q1 and Q2) alternately switch current through a center-tapped 120-V to 12-V filament transformer (T1) with its primary and secondary reversed. The 120-V output is compared with a 60 Hz reference in a closed-loop configuration that maintains a constant output at optimum efficiency.



A 60 Hz reference frequency is derived from a crystal oscillator and divider circuit, U1. An inexpensive 3.58 MHz color burst crystal provides the time base that can be accurately adjusted by C1. The 60 Hz output from U1 is applied to the exclusive-OR gate, U2, and then to the XR-2206 function generator (U3) that converts the square wave into a sine wave. U2 and U3 form a phase-locked loop that synchronizes the sine wave output of U3 with the 60 Hz square wave reference of U1. The sine wave is then inverted by op amp U4a, so that two signals 180 out of phase can be applied to U4b and U4c that drive Q1 and Q2. Due to the closed-loop configuration of the drive circuits, Q1 and Q2 conduct only during the upper half of the sine wave. Therefore, one TMOS device conducts during the first half of the sine wave and the other conducts during the second half.

5 V SUPPLY INCLUDING STABILIZED MOMENTARY BACKUP



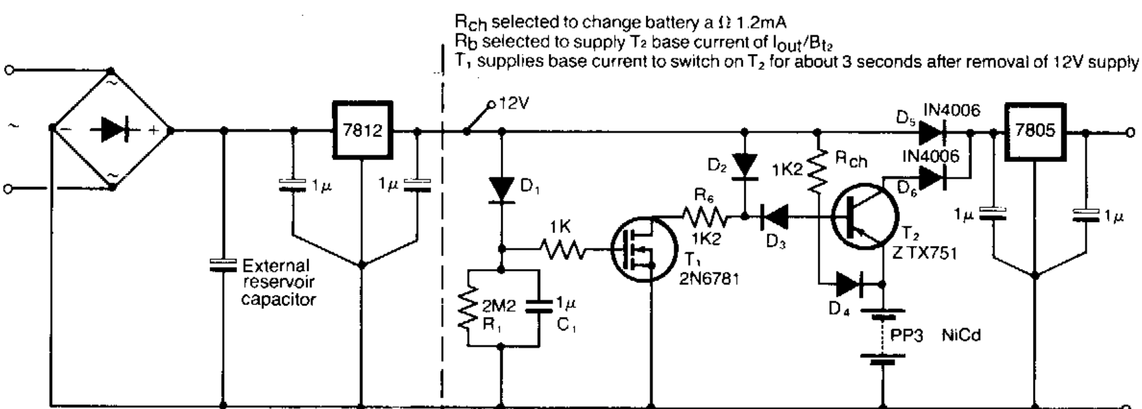


Fig. 2 12V stabilised supply (optional)

ELECTRONIC ENGINEERING

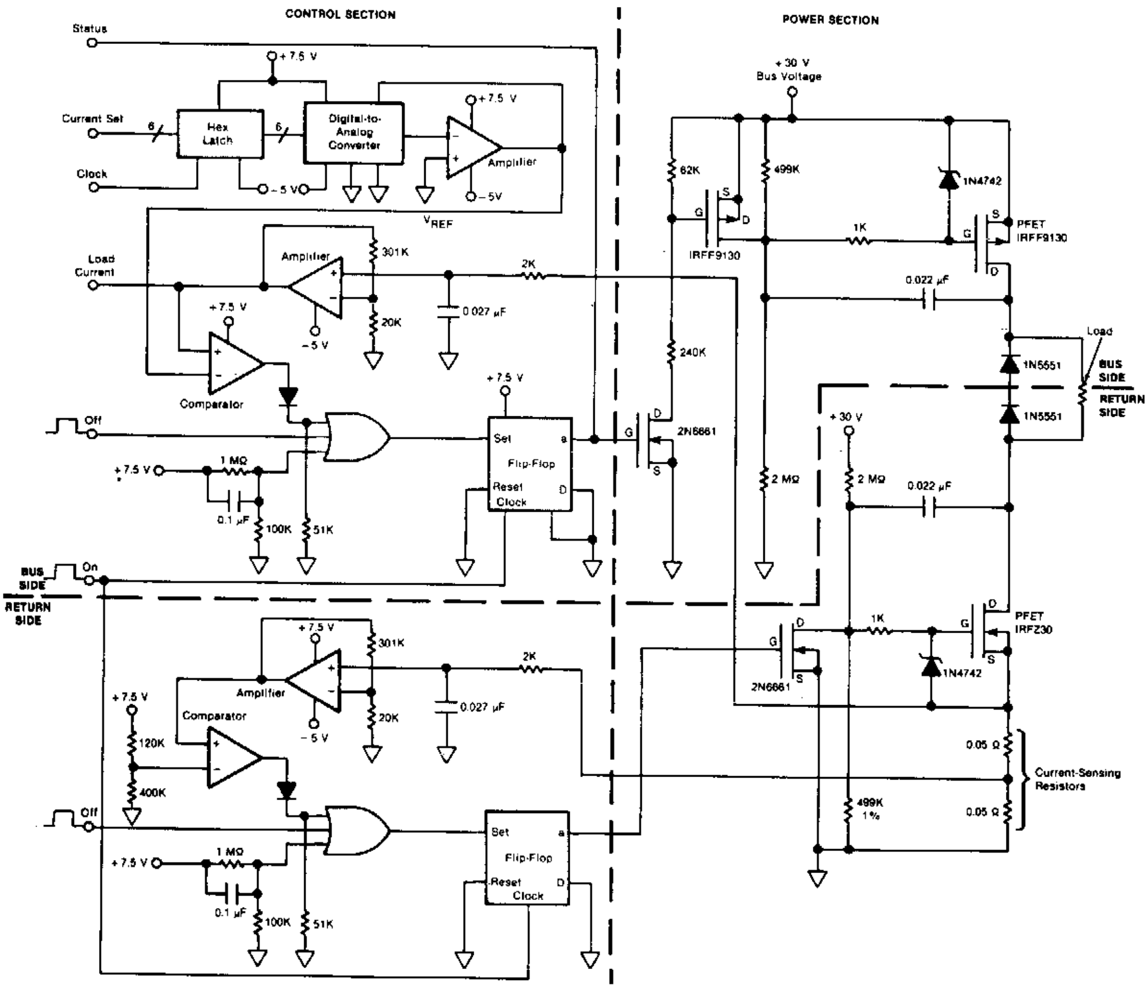
Fig. 72-3

Circuit Notes

This circuit protects microprocessor systems from "brownouts" without the expense of an uninterruptible power supply. Designed around a small 9-V nickel cadmium battery the circuit continues to provide a constant 5-V output during brownouts of up to a few seconds. Load currents of up to 500 mA may be drawn using the components shown. With this mains-derived supply present, D5 is forward biased so that the stabilized supply powers the 5-V regulator and hence the circuitry to be protected. FET T_1 is held on by D1, its drain current being provided from the dc supply via R_b and D2. Diode D3 is reverse-biased so that T2 is off, and the battery is isolated from D6. R_{CH} and D4 serve to trickle charge the battery with approximately 1.2 mA.

When the 12-V supply is removed, R1 and C1 initially keep T1 switched on. D3 is now forward biased, so that T1 drain current is drawn via R_b, D3 and T2 from the battery. This switches T2 on, allowing the load circuitry to draw current from the battery via D6 and the 5-V regulator. After a few seconds C1 has discharged (via R1) such that V_{gs} falls below the threshold value for the FET, and T1 switches off. There is then no path for T2 base current, so that it also switches off, isolating the battery.

POWER-SWITCHING CIRCUIT



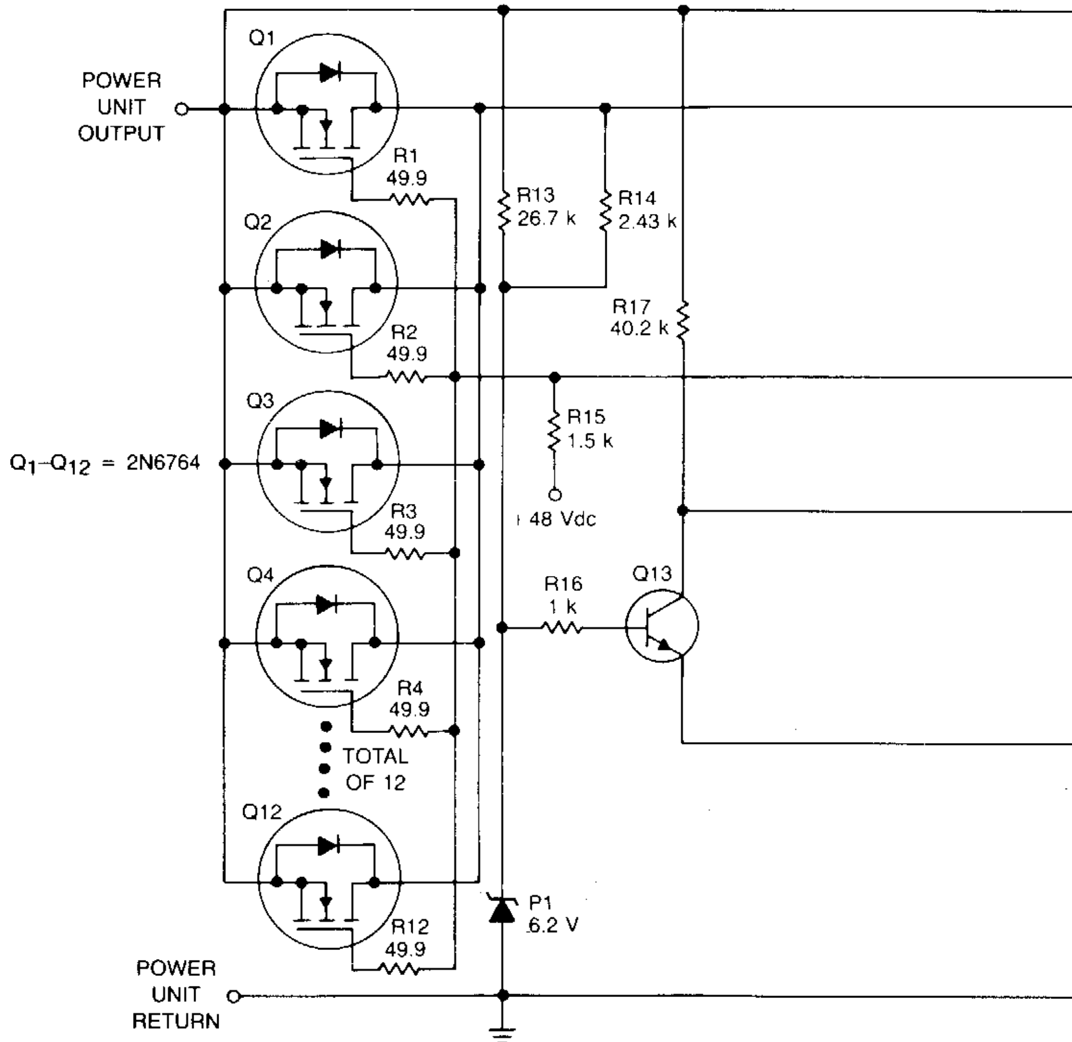
NASA

Fig. 72-4

Circuit Notes

This circuit provides on/off switching, soft starting, current monitoring, current tripping, and protection against overcurrent for a 30 Vdc power supply at normal load currents up to 2 A. The switch is turned on by an "on" command pulse; it is turned off by an "off" command pulse. An overcurrent trip can also be set on the bus side by a 6-digit binary signal, which is converted to an analog voltage and compared with the amplified voltage developed across a load-current-sensing resistor. Resistor/capacitor combinations (0.027 μF, 2 kΩ) at the inputs of the current-sensing amplifiers act as low-pass filters: this introduces a few hundred μs of delay in the response to overcurrent, thereby providing some immunity to noise. The 0.022 μF capacitors connected to the drain terminals of the PFETs provide a Miller effect, which reduces the rate of change of the drain voltage and therefore the rate of rise of current at turn-on. The soft-turn-on time depends upon the load impedance and is typically 100 to 200 ms.

RADIATION-HARDENED, 125 A LINEAR REGULATOR



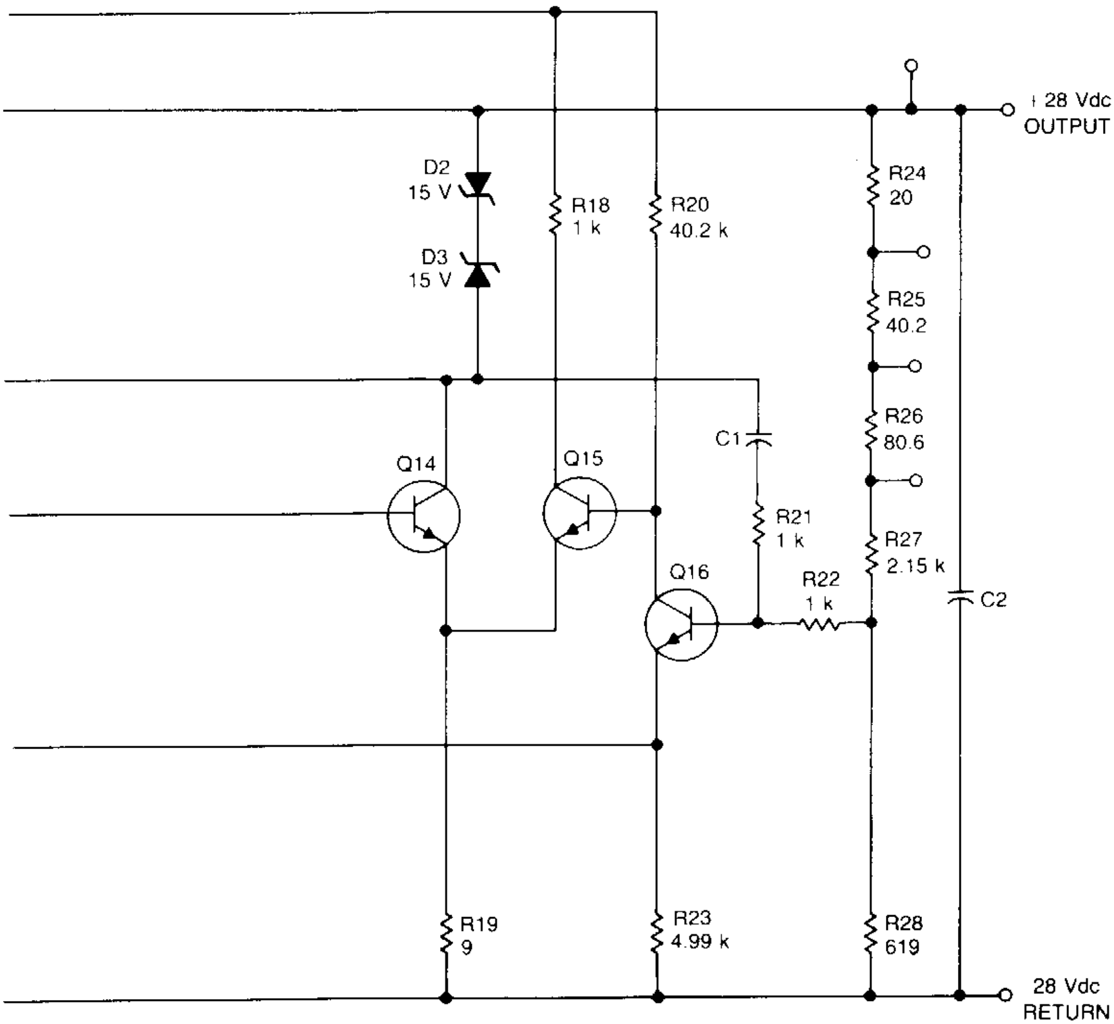
MOTOROLA

Fig. 72-5

Circuit Notes

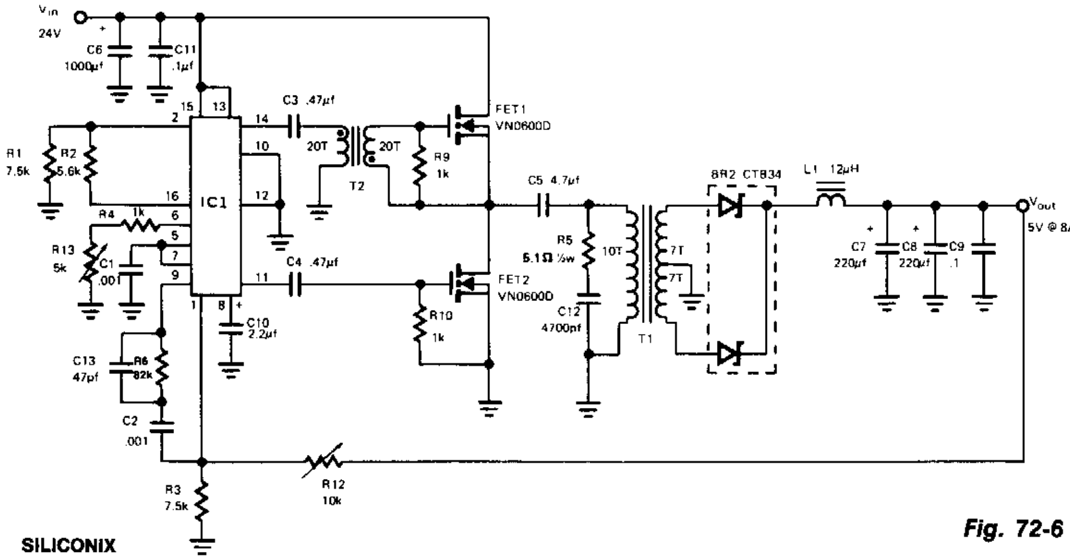
Intended for extreme temperature, radiation-hardened environments, this linear supply is capable of supplying 28 Vdc at 125 A from an ac-driven power unit.

In operation, power supply output voltage is sensed by the voltage divider consisting of R24 to R28 and fed to one input of a discrete differential amplifier composed of Q13



through Q16. The other input of the amplifier is connected to a radiation-hardened zener diode, D1. Local feedback using R21 and C1 produces gain to phase shift that are independent of individual component parameters, which provides stable operation into the required loads.

SWITCH MODE POWER SUPPLY



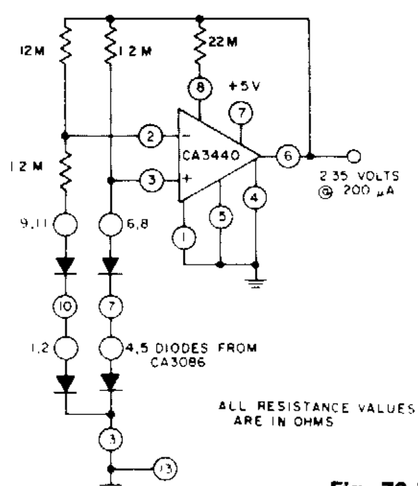
SILICONIX

Fig. 72-6

Circuit Notes

This buck-derived circuit provides up to 8 A at 5 Vdc operating off 24 to 32 Vdc. The two power MOSFETs in the circuit conduct alternately for equal periods. Switching frequency is 150 kHz, set by the PWM125 controller. Output of the two MOSFETs is transformed to a low-voltage level, then rectified. Efficiency of the circuit is 75% when operated in a 22 to 32 V range. Efficiency approaches 90% with higher voltage inputs.

MICROPOWER BANDGAP REFERENCE SUPPLY



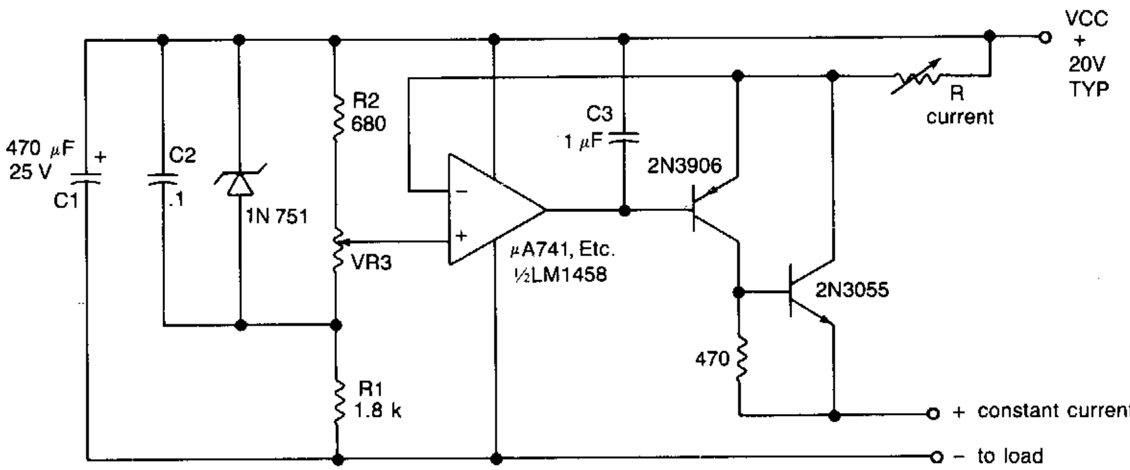
GENERAL ELECTRIC/RCA

Fig. 72-7

Circuit Notes

The circuit uses a CA3440 BiMOS op amp and CA3086 transistor array. The no-load current from 5-volt supply is 1.5 μA. Load current can go as high as 200 μA and still maintain output voltage regulation within 0.05%.

VARIABLE CURRENT SOURCE, 100 mA TO 2 AMP



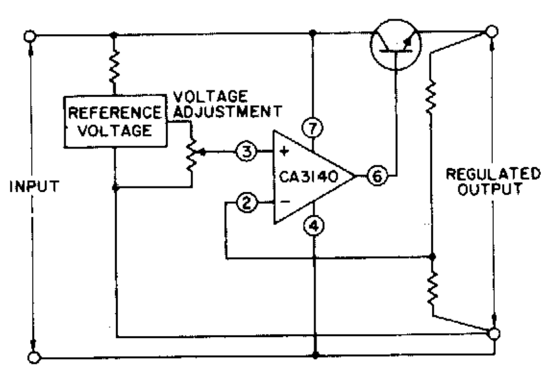
WILLIAM SHEETS

Fig. 72-8

Circuit Notes

The output current is set by the resistor R in the collector of Tr2, which may be varied to offer a range of output currents from 100 mA to 2 A with fine control by means of VR3 which varies the reference voltage to the non-inverting input of the op amp. The feedback path from the output to the inverting input of the op amp maintains a constant voltage across R, equal to $(V_{CC}\text{-}V_{IN})$ and hence a constant current to the load given by $(V_{CC}\text{-}V_{IN})/R$.

BASIC SINGLE-SUPPLY VOLTAGE REGULATOR



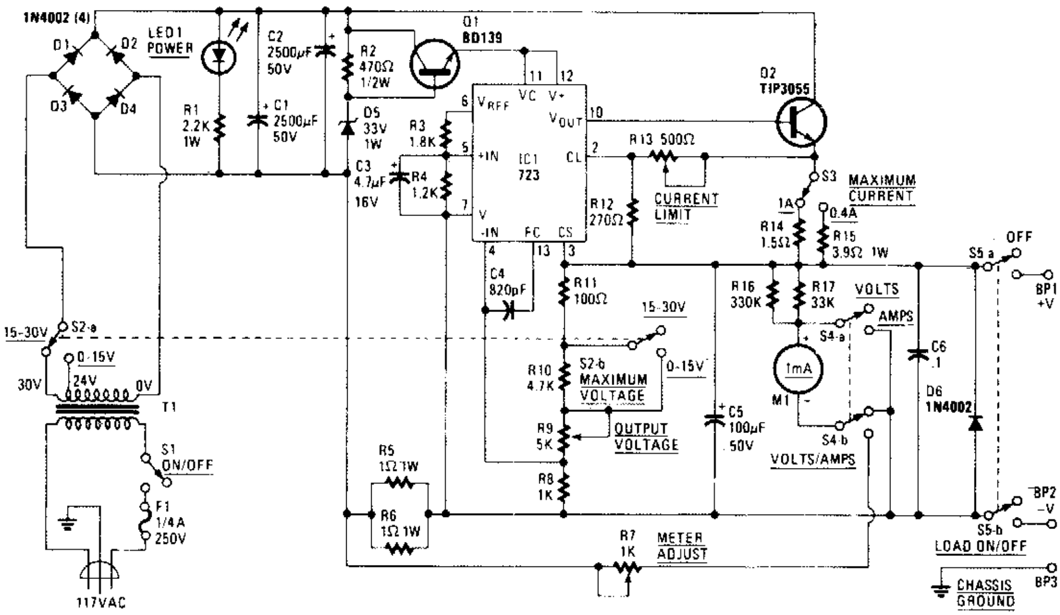
Circuit Notes

The circuit uses a CA3140 BiMOS op amp capable of supplying a regulated output that can be adjusted from essentially 0 to 24 volts.

GENERAL ELECTRIC/RCA

Fig. 72-9

BENCH TOP POWER SUPPLY

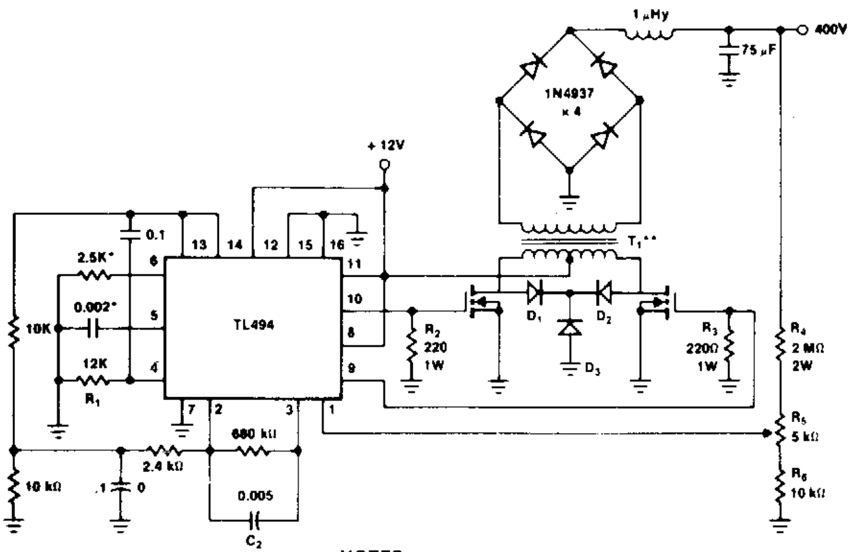


RADIO-ELECTRONICS *Fig. 72-10*

Circuit Notes

A tapped transformer drives a diode bridge (D1-D4) and two 2500 μF filter capacitors (C1 and C2), that provide a no-load voltage of 37 or 47 volts, depending upon the position of switch S2a. The unregulated dc is then fed to a pre-regulator stage composed of Q1 and D5. Those components protect IC1 (the 723) from an over-voltage condition; the 723 can't handle more than 40 volts. The LED (LED1) and its 2.2 k current-limiting resistor (R1) provide on/off indication. The current through the LED varies slightly according to the transformer tap selected, but that's of no real consequence. The series-pass transistor in IC1 drives voltage-follower Q2, providing current amplification. The transistor can handle lots of power. It has a maximum collector current of 15 amps and a maximum V_{CE} of 70 V, both of which are more than adequate for our supply.

400-VOLT, 60-WATT PUSH-PULL POWER SUPPLY



NOTES
UNLESS OTHERWISE NOTED,
ALL RESISTORS 5%, ¼ W
ALL CAPACITOR VALUES IN MICROFARADS, 25V
Q_1 & Q_2: VN64GA ON HEAT SINK
D_1 & D_2: 1N4934
D_3: 33V, 3W ZENER
T_1: PRI: 12T, CT, NO 18 AWG
SEC: 275T, NO 24 AWG
CORE: IND GEN 8231-1

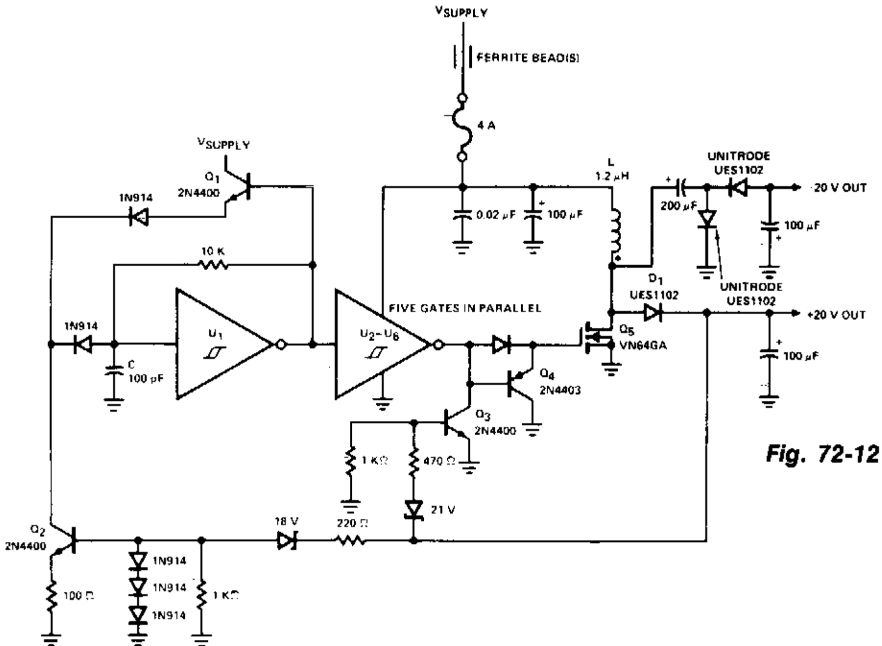
SILICONIX, INC.

Fig. 72-11

Circuit Notes

The design delivers a regulated 400-V, 60-W output. The TL494 switching regulator governs the operating frequency and regulates output voltage. R1 and C1 determine switching frequency, which is approximately 0.5RC—100 kHz for the values shown. The TL494 directly drives the FET's gates with a voltage-controlled, pulse-width-modulated signal. After full-wave rectification, the output waveform is filtered by a choke-input arrangement. The 1 μH, 75 μF filter accomplishes the job nicely at 100 kHz. A feedback scheme using R4, R5 and R6 provides for output-voltage regulation adjustment, with loop compensation handled by C2. Diodes D1 and D2 provide isolation and steering for the 33-V zener transient clamp, D3. Output regulation is typically 1.25% from no-load to the full 60-W design rating. Regulation is essentially determined by the TL494. Output noise and ripple consists mainly of positive and negative 0.8-V spikes occurring when the output stage switches.

500 kHz SWITCHING INVERTER FOR 12 V SYSTEMS



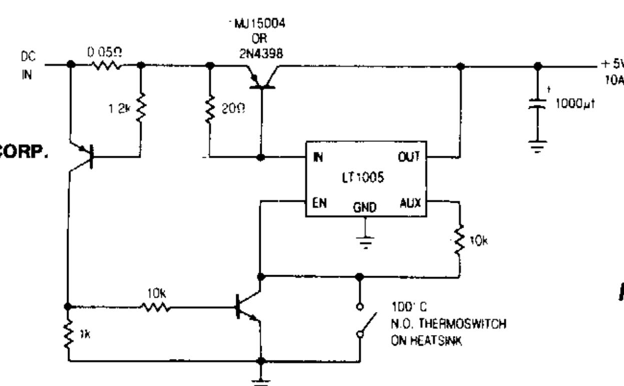
SILICONIX, INC.

Fig. 72-12

Circuit Notes

This PWM control circuit provides the control pulse to the DMOS Power Switch in the flyback circuit. The output of the PWM is a pulse whose width is proportional to the input control voltage and whose repetition rate is determined by an external clock signal. To provide the control input to the PWM and to prevent the output voltage from soaring or sagging as the load changes the error amplifier and reference voltage complete the design. They act as the feedback loop in this control circuit much like that of a servo control system.

10 AMP-REGULATOR WITH CURRENT AND THERMAL PROTECTION



LINEAR TECHNOLOGY CORP.

Fig. 72-13

BIPOLAR POWER SUPPLY FOR BATTERY INSTRUMENTS

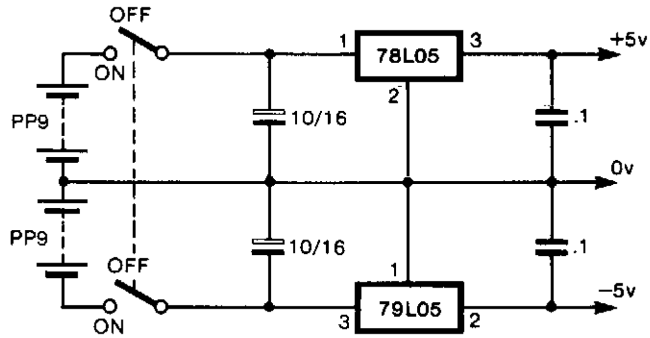


Fig. 1

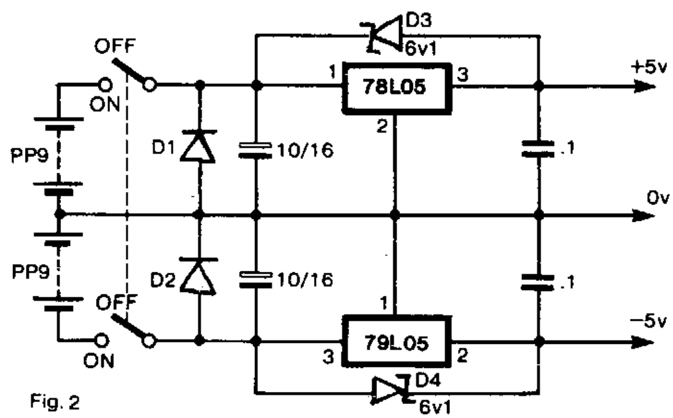


Fig. 2

ELECTRONIC ENGINEERING

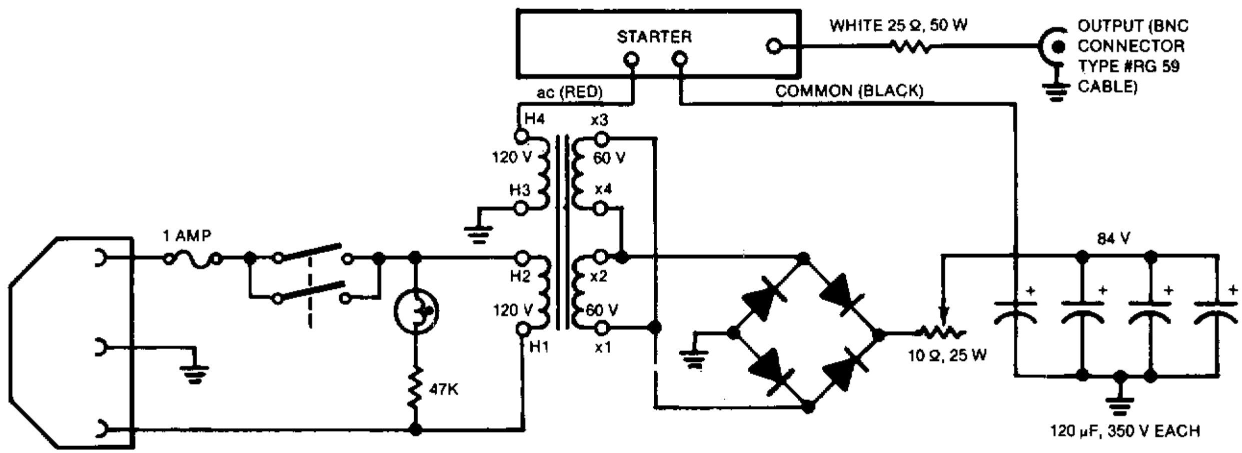
Fig. 72-14

Circuit Notes

To generate regulated ±5-V supplies from a pair of dry batteries, the circuit of Fig. 1 is commonly used. In order to give protection from inadvertent reverse connection of a battery, a diode in series with each battery would produce an unacceptable voltage drop. The more effective approach is to fit diodes D1 and D2 as shown in Fig. 2, in parallel with each battery.

When the supply is switched off, there is the risk of a reverse bias being applied across the regulators, if there is significant inductance or capacitance in the load circuit. Diodes across the regulators prevent damage. When the power supply is switched on, the two switches do not act in unison. There is a probability that one or the other regulators will be latched hard off by the other. To prevent this, D3 and D4 are Zener diodes so that ±5-V rails are pulled up by the batteries until the regulators establish the correct levels.

POWER SUPPLY FOR 25-WATT ARC LAMP



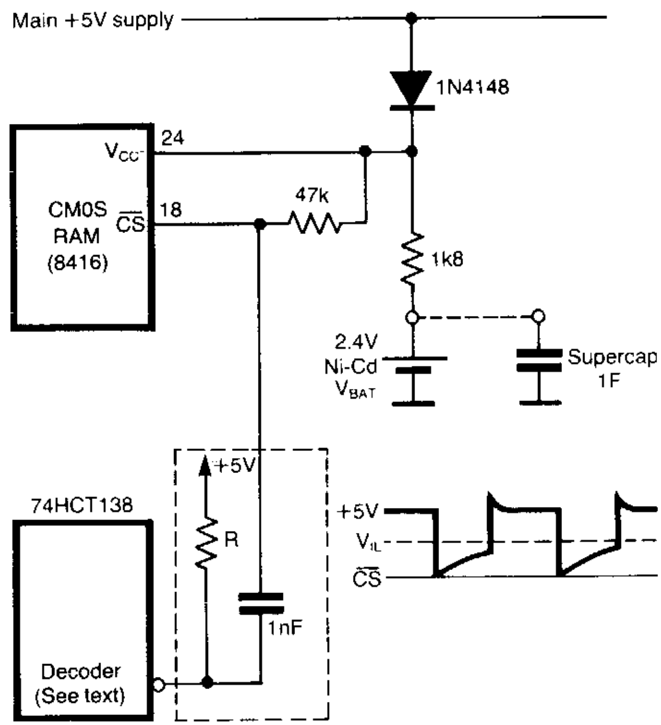
NASA

Fig. 72-15

Circuit Notes

A dual-voltage circuitry both strikes and maintains the arc. The lamps require a starting voltage in excess of 1,000 volts. Once stabilized, the voltage drop across the lamp is near 20 volts. Power supply consists of two main sections. The first section, the low-voltage power supply section, is an 84-volt direct-current supply. This supply powers the stabilized arc. Current is limited by the 10 ohm adjustable and 25 ohm fixed resistance. The second section, the high-voltage starter circuit, is a Cockroft-Walton voltage multiplier. With no load, the output voltage is 2,036 volts. However, when the arc is established, the heavy current drain maintains a forward bias on all of the diodes, and the circuit becomes a straight path with a voltage drop of 7.2 volts. The small value of the capacitors used in the multiplier guarantees that the diodes will be forward-biased once the arc is established.

STAND-BY POWER FOR NON-VOLATILE CMOS RAMs



ELECTRONIC ENGINEERING *Fig. 72-16*

Circuit Notes

To prevent loss of data when a CMOS RAM is switched from normal operation ($V_{CC}=5$ volts) to stand-by mode ($V_{CC}=V_{BAT}$) it must be ensured that the CS pin goes near the V_{CC} rail at all times. Ac coupling to the chip select is made through capacitor C, breaking the dc current path between V_{CC} (and hence V_{BAT}) and the decoder output. So, whatever the impedance state of the decoder in power down, the battery will provide current only for the RAM, low enough to keep the voltage at CS near to V_{CC}.

HV REGULATOR WITH FOLDBACK CURRENT LIMITING

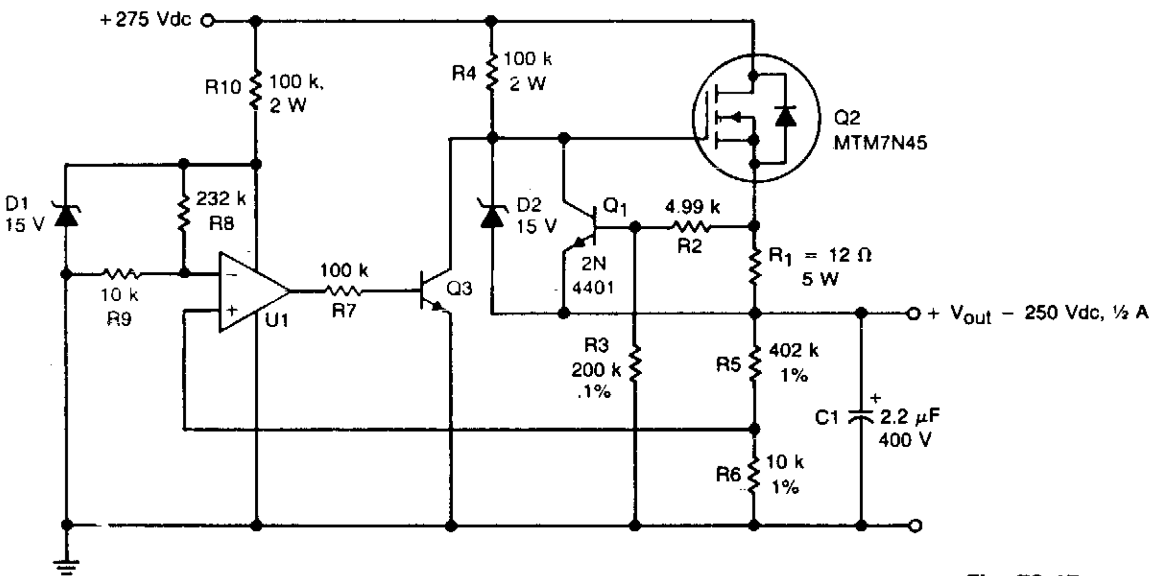


Fig. 72-17

MOTOROLA

Circuit Notes

A TMOS MTM7N45 (Q2) is used as a series pass element in a linear high voltage supply that accepts +275-V unregulated and produces 250 V regulated with foldback current limiting.

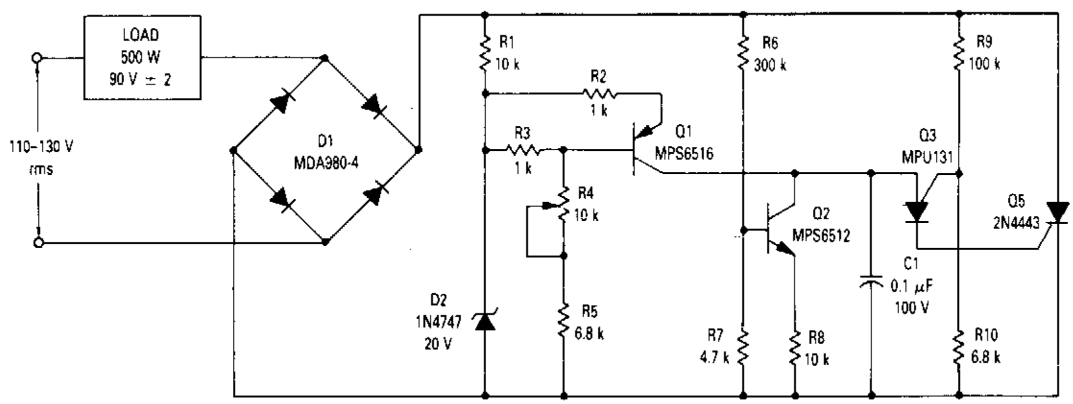
A 15-V zener, D1, provides the dc reference for operational amplifier U1, whose other input is obtained from a fraction of the output voltage. U1 drives Q3, which drives the gate of Q2. Foldback current limiting is achieved by R1, R2, R3, R4, Q1, and D2. The formula to establish the current "knee" for limiting is:

$$I_{KNEE} = \frac{V_{OUT}(R2/R2 + R3) + 0.5\ V}{R1}$$

Short circuit current is:

$$I_{SC} = \frac{0.5\ V}{R1}$$

90 V rms VOLTAGE REGULATOR USING A PUT



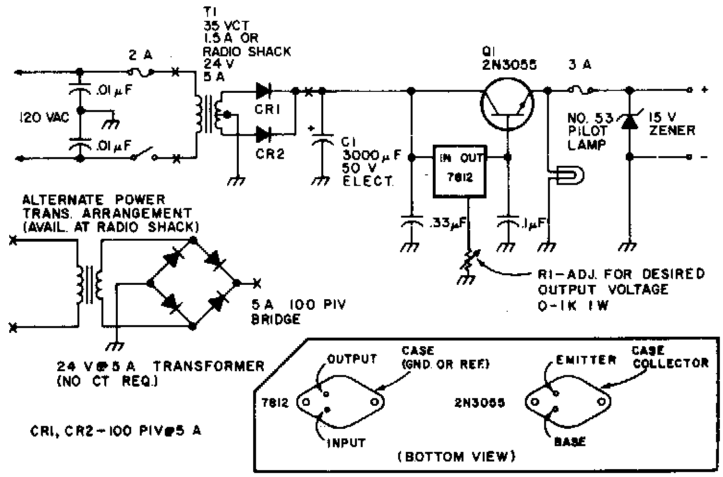
MOTOROLA

Fig. 72-18

Circuit Notes

The circuit is an open loop rms voltage regulator that will provide 500 watts of power at 90 V rms with good regulation for an input voltage range of 110-130 V rms. With the input voltage applied, capacitor C1 charges until the firing point of Q3 is reached causing it to fire. This turns Q5 on which allows current to flow through the load. As the input voltage increases, the voltage across R10 increases which increases the firing point of Q3. This delays the firing of Q3 because C1 now has to charge to a higher voltage before the peak-point voltage is reached. Thus the output voltage is held fairly constant by delaying the firing of Q5 as the input voltage increases. For a decrease in the input voltage, the reverse occurs.

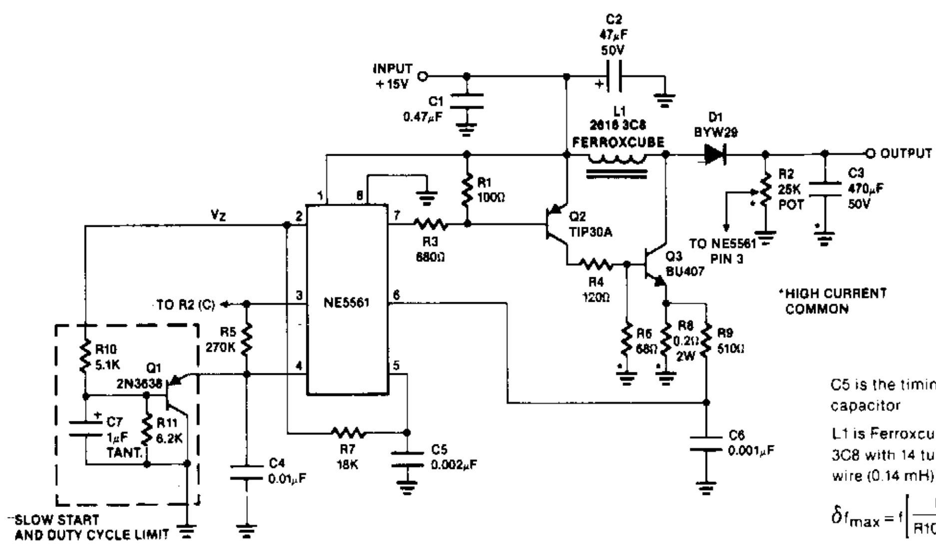
12-14 V REGULATED 3 A POWER SUPPLY



AMATEUR RADIO

Fig. 72-19

DC-TO-DC SMPS VARIABLE 18 V TO 30 V OUT AT 0.2 A



C5 is the timing capacitor

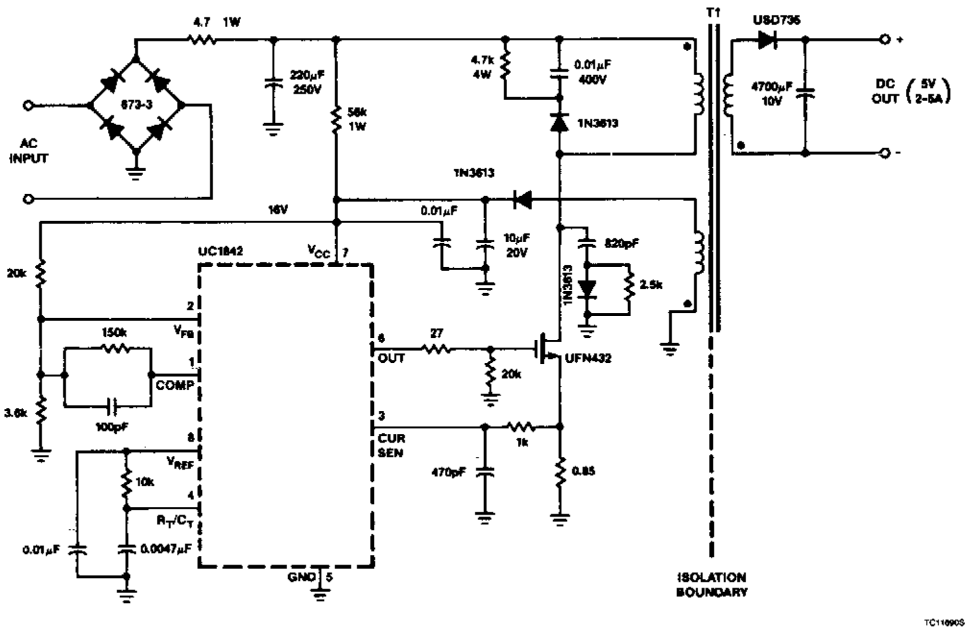
L1 is Ferroxcube 2616 3C8 with 14 turns of #18 enameled wire (0.14 mH)

$$\delta f_{max} = f\left[\frac{R11}{R10 + R11}\ (V_Z) + V_{BE}\right]$$

SIGNETICS

Fig. 72-20

OFF-LINE FLYBACK REGULATOR



NOTES:
T1: Coilcraft E-4140-B
Primary - 97 turns
single AWG 24
Secondary - 4 turns
4 parallel AWG 22
Control - 9 turns
3 parallel AWG 28

SPECIFICATIONS

Input line voltage:	$90V_{AC}$ to $130V_{AC}$	Efficiency @ 25 W,	
Input frequency:	50 or 60Hz	$V_{IN} = 90V_{AC}$:	70%
Switching frequency:	40kHz± 10%	$V_{IN} = 130V_{AC}$:	65%
Output power:	25W maximum	Output short-circuit current:	2.5A average
Output voltage:	5V± 5%		
Output current:	2 to 5A		
Line regulation:	0.01%/V		
Load regulation:	8%/A*		

SIGNETICS

Fig. 72-21

Circuit Notes

This circuit uses a low-cost feedback scheme in which the dc voltage developed from the primary-side control winding is sensed by the UC1842 error amplifier. Load regulation is therefore dependent on the coupling between secondary and control windings, and on transformer leakage inductance. For applications requiring better load regulation, a UC1901 Isolated Feedback Generator can be used to directly sense the output voltage.

SCR PREREGULATOR FITS ANY POWER SUPPLY

Fig. 1

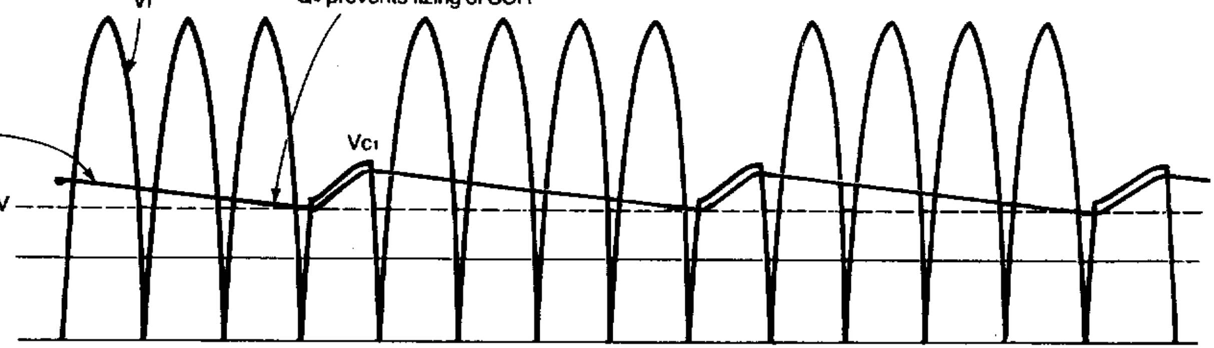
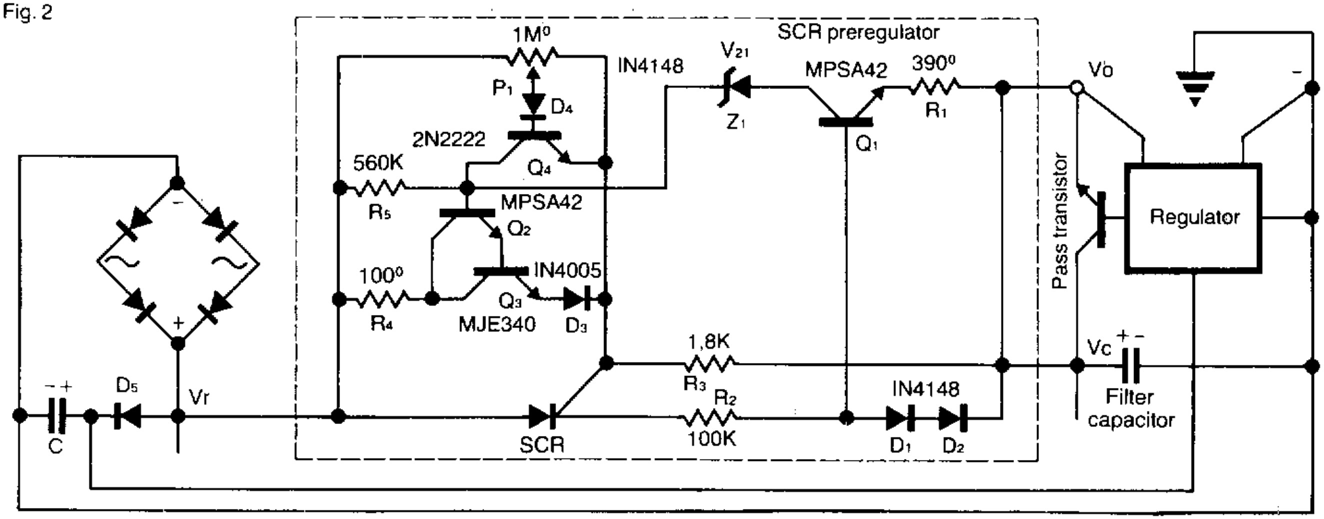


Fig. 2



ELECTRONIC ENGINEERING

Fig. 72-22

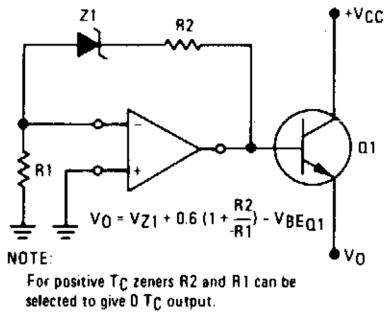
Circuit Notes

This SCR pre-regulator keeps the filter capacitor V_c, in a variable output power supply, a few volts above the output voltage V_o. The benefits include: less heat dissipated by the pass transistor and therefore small heatsink, cooler operation and higher efficiency, especially at low output voltages.

Q1, R1, R2, D1 and D2 form a constant current source for zener Z1, so that the contribution to the output current is always a few mA (2 - 3 mA).

The Darlington pair Q2, Q3 keeps the SCR off. The voltage V_c decreases until $V_c = V_o = V$ at which point the Darlington pair fires the SCR, charging the filter capacitor to a higher voltage V_c^1 in less than half the period of the input voltage. The component values, shown are for a 0 - 250-V, 3-A power supply.

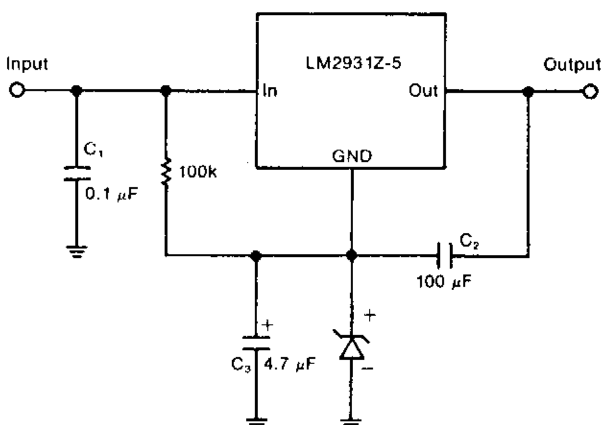
VOLTAGE REGULATOR



MOTOROLA, INC.

Fig. 72-23

ZENER DIODE INCREASE FIXED PNP REGULATOR'S OUTPUT VOLTAGE RATINGS



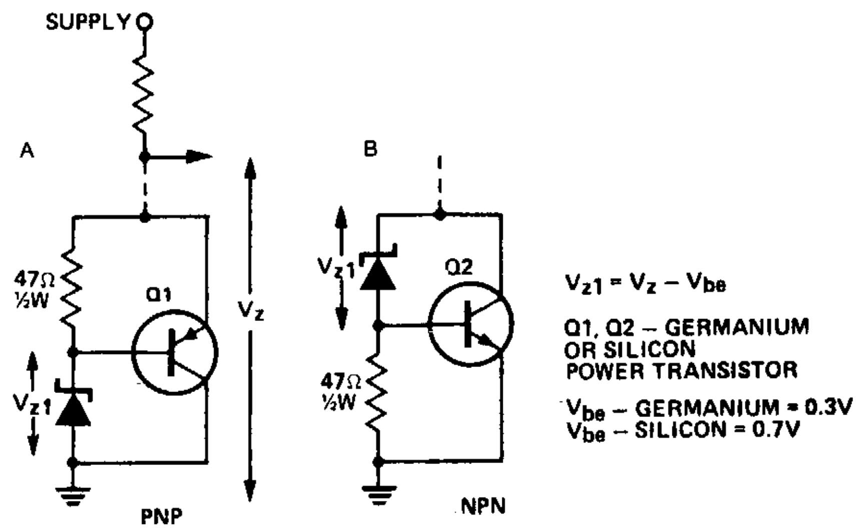
ELECTRONIC DESIGN

Fig. 72-24

Circuit Notes

A zener diode in the ground lead of a fixed pnp regulator varies the voltage output of that device without a significant sacrifice in regulation. The technique also allows the regulator to operate with output voltages beyond its rated limit.

INCREASING THE POWER RATING OF ZENER DIODES



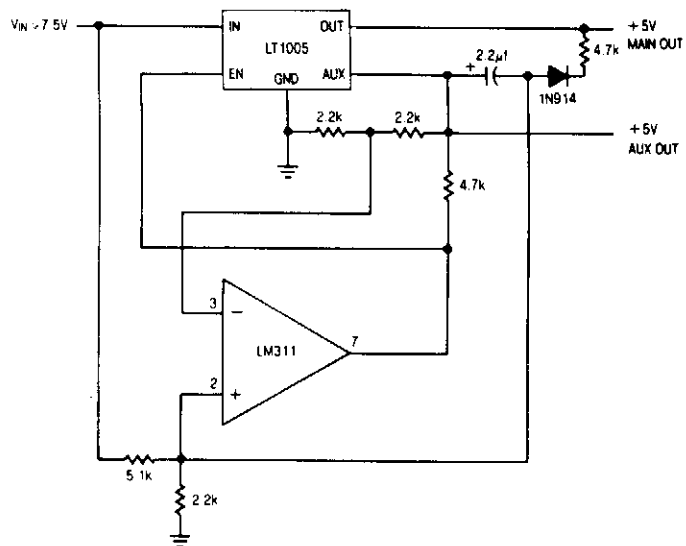
ELECTRONICS TODAY INTERNATIONAL

Fig. 72-25

Circuit Notes

A power transistor can be used to provide a high powered zener voltage from a low wattage zener. A 400 mW zener can be used where a 10 watt zener is required or a 1 W zener can be used where a 50 to 80 watt zener is required by using appropriate transistors for Q1 and Q2 in the circuits shown. Where low rating is required, Q1 would be an ASZ 15 (germanium) or an AY9140 (silicon). Q2 could be a 2N2955 (silicon). For higher powers, Q1 should be an ASZ18 (germanium) or a 2N2955 (silicon) and Q2 a 2N3055 (silicon) or an AY8149 (silicon). A heatsink on the transistor is required. The circuit in *A* has the advantage that power transistors can be bolted directly on to a chassis which may serve as a heatsink.

MEMORY SAVE ON POWER-DOWN



LINEAR TECHNOLOGY *Fig. 72-26*

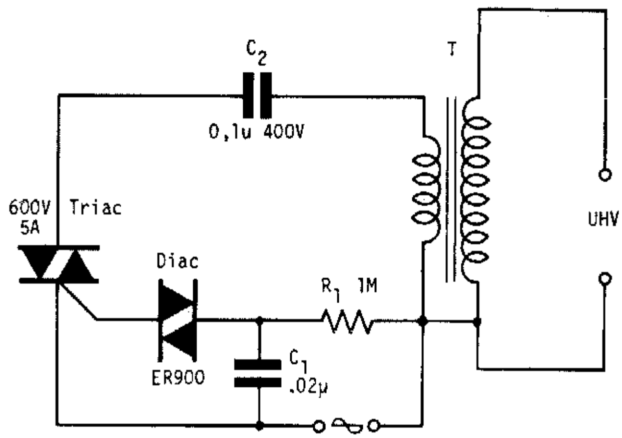
Circuit Notes

The auxiliary output powers the memory, while the main output powers the system and is connected to the memory store pin. When power goes down, the main output goes low, commanding the memory to store. The auxiliary output then drops out.

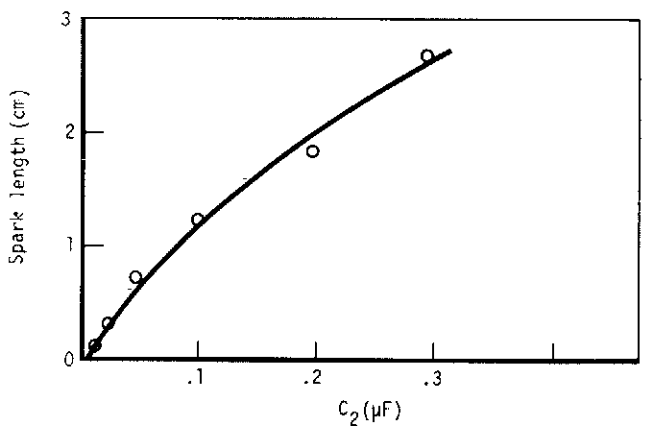
73
Power Supply Circuits (High Voltage)

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW COST ULTRA HIGH VOLTAGE GENERATOR



A



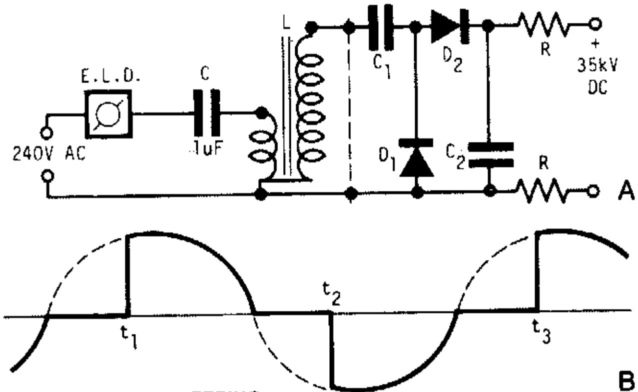
B

ELECTRONIC ENGINEERING *Fig. 73-1*

Circuit Notes

By repetitively charging and discharging a capacitor through the primary of an induction coil with a high voltage, an ultra high emf is induced in the secondary. Switching is performed by the triac, triggered by the disc at times set by C1 and R1. With a 12 V car ignition coil for example, the length of sparkgap obtained is 12 mm of air for C2 = 0.1 μF. If the dielectric strength of air is assumed to be 3 kV/mm, this spark-gap length corresponds to 36 kV. From the curve shown in Fig. B, care must be taken in keeping the value of C2 below 1 μF as the coil is liable to be seriously damaged at this value of C2. Power consumption is only about one watt.

SIMPLE HIGH-VOLTAGE SUPPLY



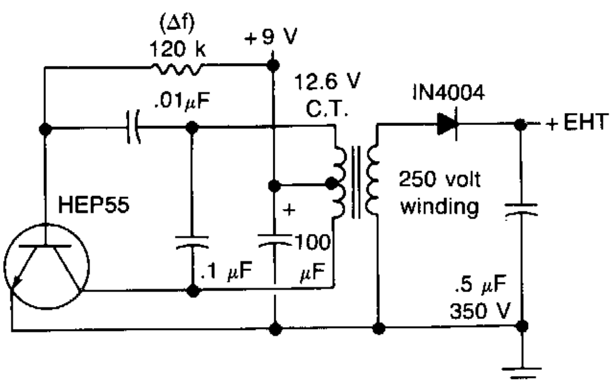
ELECTRONIC ENGINEERING

Fig. 73-2

Circuit Notes

A light dimmer, a 1 μF capacitor and a 12 V car ignition coil form the simple line powered HV generator. The current in the dimmer is shown in Fig. B. At times t_1, t_2, . . . , set by the dimmer switch, the inner triac of the dimmer switches on, and a very high and very fast current pulse charges the capacitor through the primary of the induction coil. Then at a rate of 120 times per second for a 60 Hz line, a very high voltage pulse appears at the secondary of the coil. To obtain an HV dc output, use a voltage doubler. D1 and D2 are selenium rectifiers (TV 18 Siemens or ITT) used for the supply of television sets. High value output shock protection resistors, R, are recommended when suitable.

HIGH VOLTAGE GEIGER COUNTER SUPPLY



POPULAR ELECTRONICS

Fig. 73-3

Circuit Notes

This circuit will generate about 300 volts dc—at a very low current, but enough for a GM tube.

HIGH VOLTAGE SUPPLY

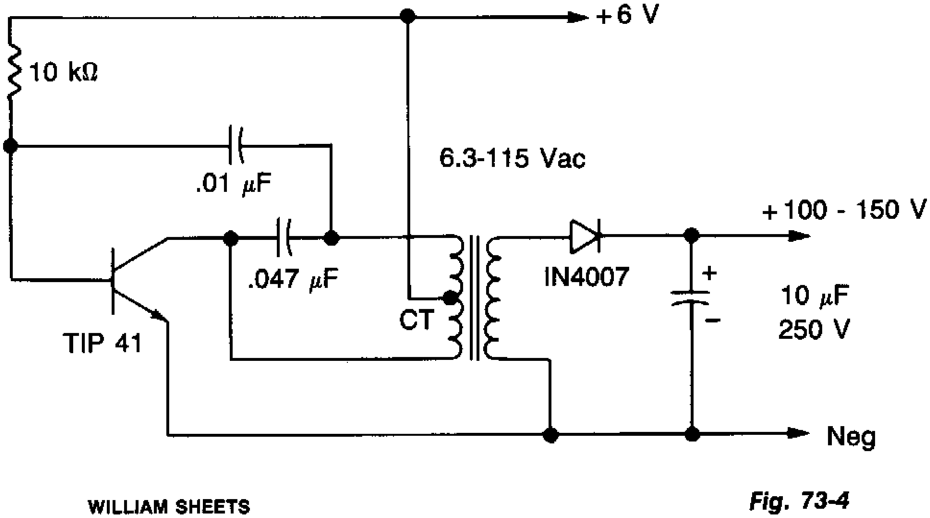


Fig. 73-4

Circuit Notes

A 6 V battery can provide 100-150 Vdc center-tapped at a high internal impedance (not dangerous though it can inflict an unpleasant jolt). A 6.3 V transformer is connected "in reverse" with a transistor used in a Hartley oscillator configuration. The frequency of operation may be controlled by varying the value of the 10 K ohm resistor. The 10 μF capacitor must have a working voltage of at least 250 Vdc.

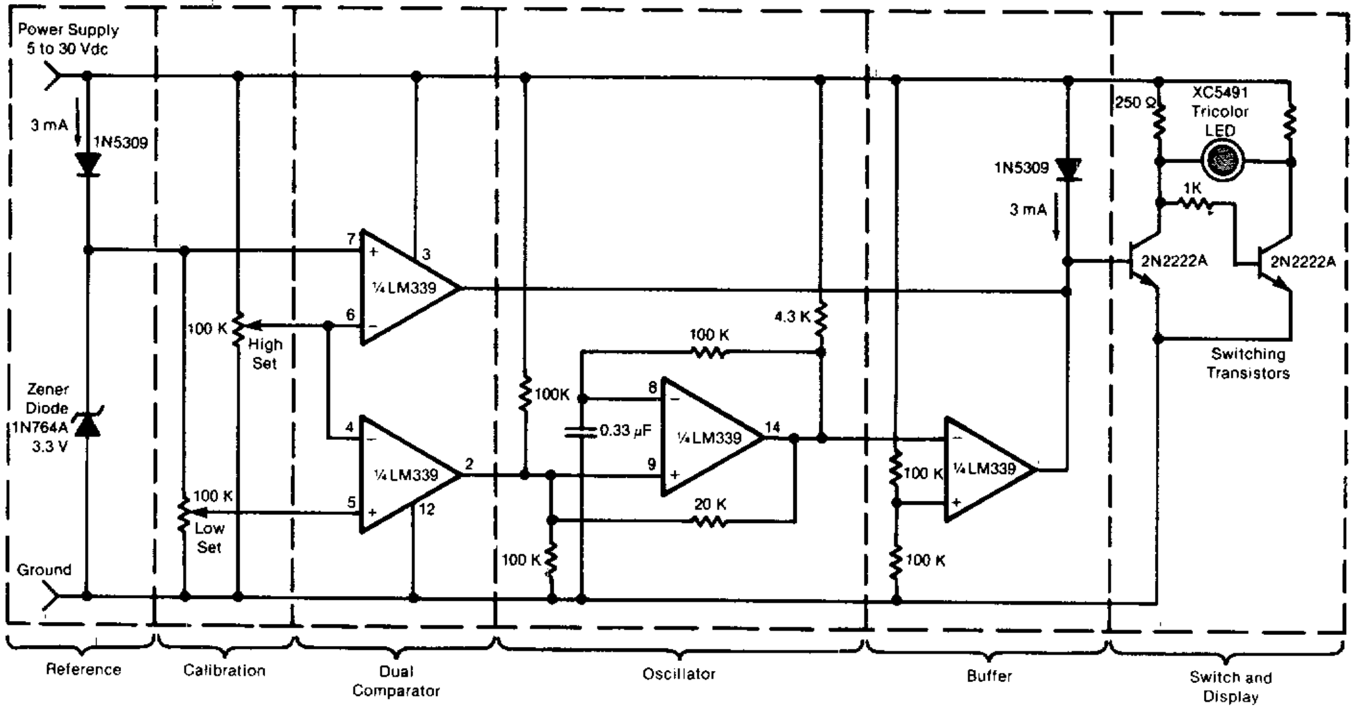
74

Power Supply Monitors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Power Supply Monitor
Low-Volts Alarm
Microprocessor Power Supply Watchdog
Overvoltage Protection Circuit (SCR Crowbar)
Power Supply Protection Circuit

POWER SUPPLY MONITOR



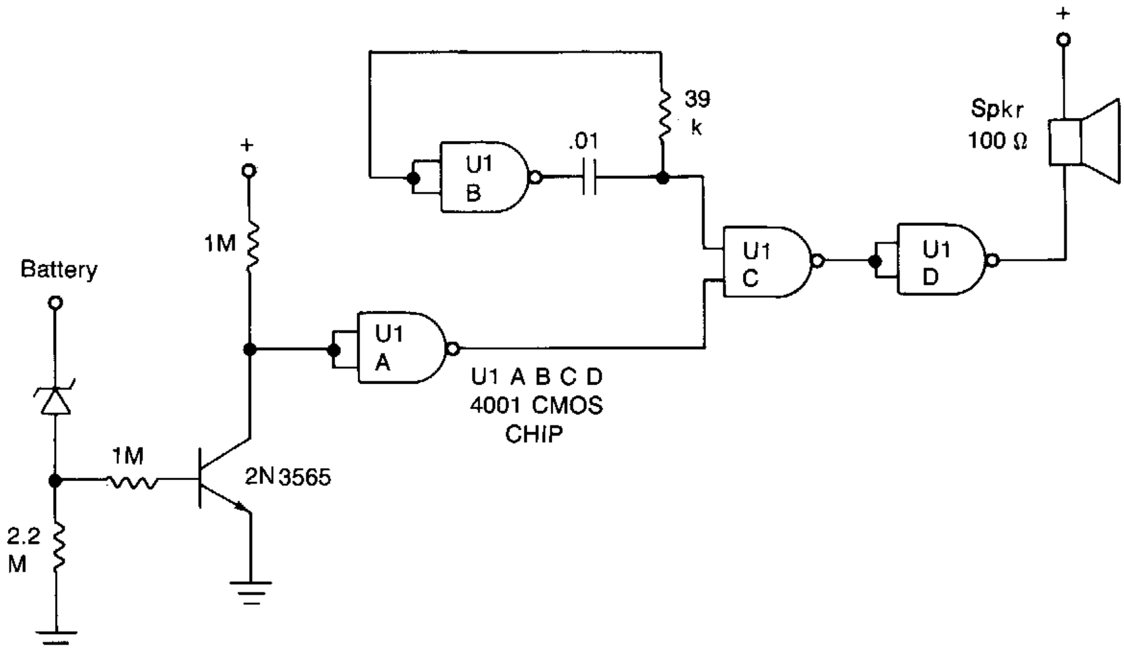
NASA

Fig. 74-1

Circuit Notes

This circuit uses a tricolor LED display to indicate acceptable and unacceptable output voltages. One to set the upper voltage limit, the other, the lower voltage limit. When the monitored voltage is above the set maximum, the LED display turns red. Yellow turns on for voltages below the set minimum, and green turns on for voltages between the high and the low settings. The circuit does not need a separate power supply. It is powered by the voltage it monitors. The circuit can be adapted to monitor voltage differences between two power supplies. Should the monitored voltages differ by more than a set value, a visual or an audible alarm would warn the operator about the difference. The circuit can also be modified for remote monitoring and the use of a separate power supply.

LOW-VOLTS ALARM



WILLIAM SHEETS

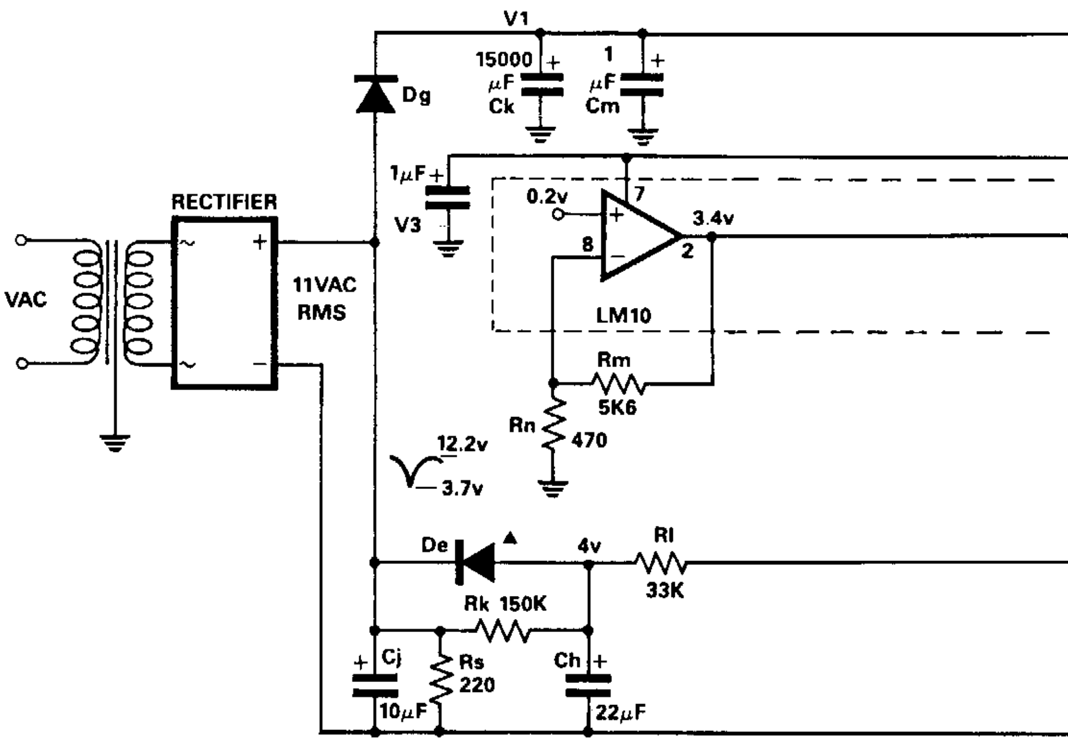
Fig. 74-2

Circuit Notes

This inexpensive dc supply-voltage monitor sounds a warning when the voltage falls below a preset value. It is ideal for monitoring rechargeable batteries since it draws only a few microamperes when not sounding. The voltage at which the alarm sounds is determined by the zener diode. When the voltage falls below the zener voltage, the alarm sounds. The alarm tone is determined by the RC time constant of the 39 k resistor and 0.01 mf capacitor.

MICROPROCESSOR POWER SUPPLY WATCHDOG

Fig 1

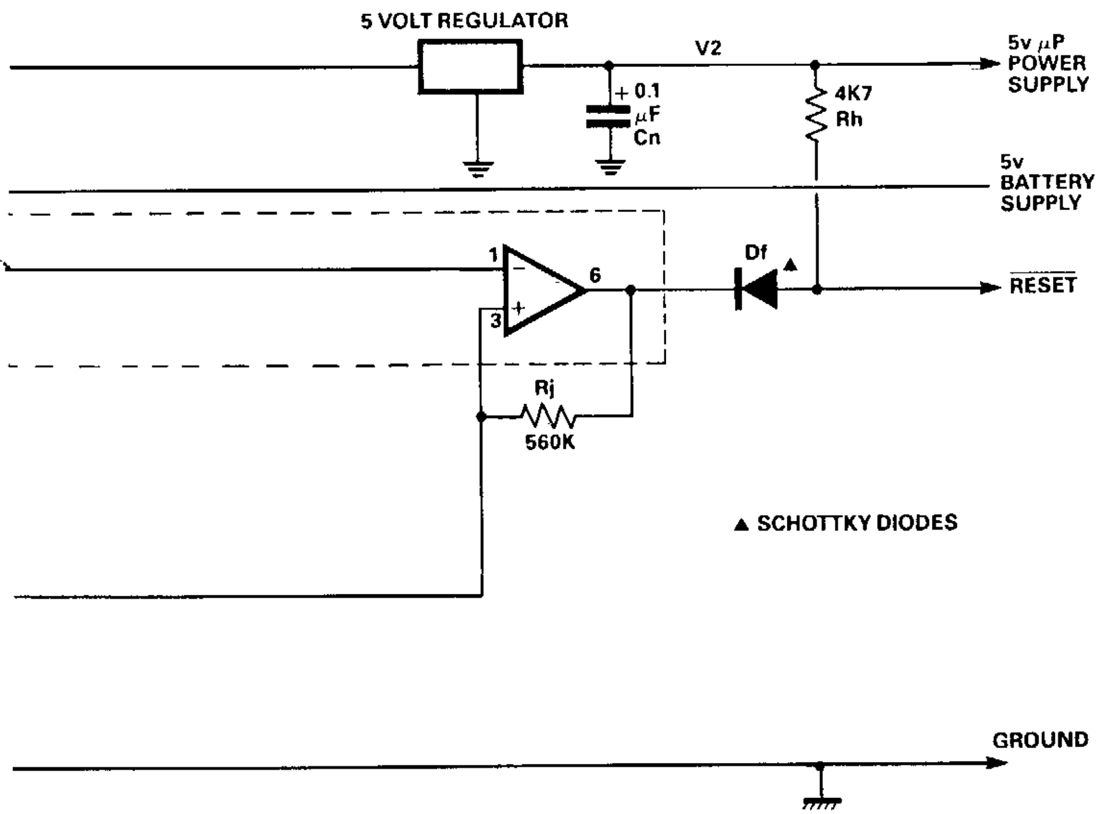


ELECTRONIC ENGINEERING

Fig. 74-3

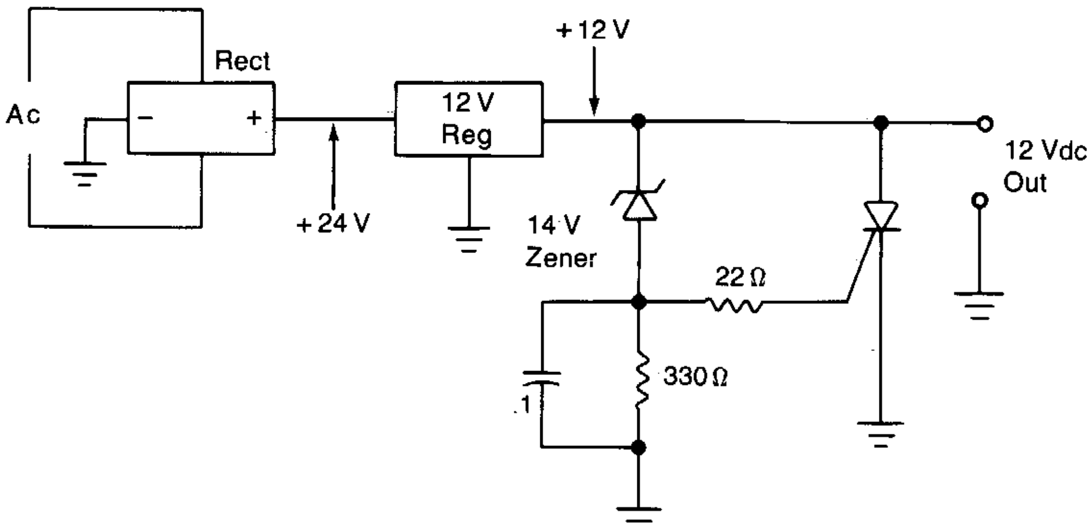
Circuit Notes

The circuit monitors the input to the microprocessor 5 V regulated supply for voltage drops and initiates a reset sequence before supply regulation is lost. In operation, the resistor capacitor combination Rs and Cj form a short time constant smoothing network for the output of the fullwave bridge rectifier. An approximately triangular, voltage waveform appears across C and Rs and it is the minimum excursion of this that initiates



the reset. Diode Dg prevents charge sharing between capacitors Cj and Ck. Resistors Rn and Rm form a feedback network around the voltage reference section of the LM10C, setting a threshold voltage of 3.4 volts. The threshold voltage is set at 90% of the minimum voltage of the triangular waveform. When the triangular wave trough, at the comparator's non-inverting input, dips below the threshold, the comparator output is driven low. This presents a reset to the microprocessor. Capacitor Ch is charged slowly through resistor Rk and discharged rapidly through diode De.

OVERVOLTAGE PROTECTION CIRCUIT (SCR CROWBAR)



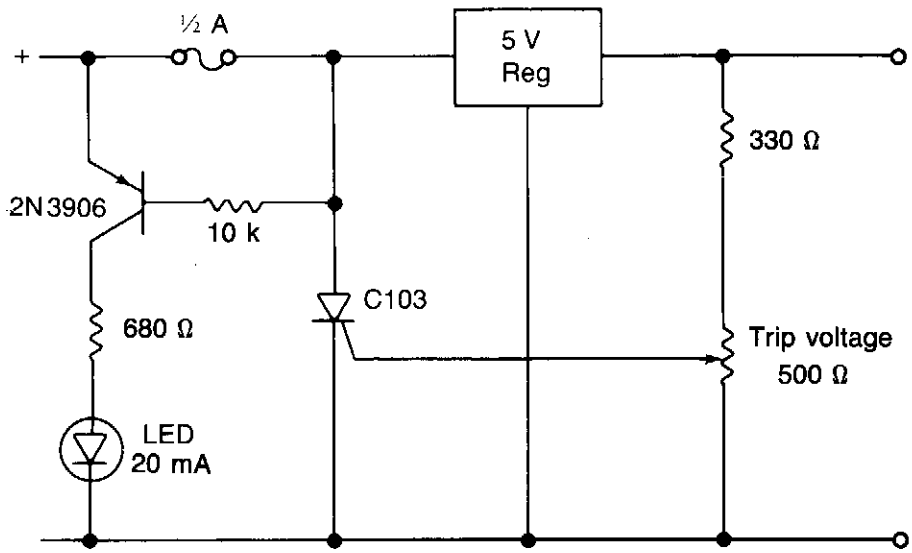
WILLIAM SHEETS

Fig. 74-4

Circuit Notes

The silicon controlled rectifier (SCR) is rated to handle at least the current of the power supply. It is connected in parallel across the 12 V dc output lines, but remains inert until a voltage appears at the gate terminal. This triggering voltage is supplied by the zener diode. At potentials less than 14 V the zener will not conduct current. But, at potentials greater than 14 Vdc the zener conducts and creates a voltage drop across the 330 ohm resistor that will fire the SCR. When the SCR turns on, the output lines of the power supply are shorted to ground. This will blow the primary fuse or burn out the transformer if there is no primary fuse.

POWER SUPPLY PROTECTION CIRCUIT



WILLIAM SHEETS

Fig. 74-5

Circuit Notes

When using a regulated supply to reduce a supply voltage there is always the danger of component failure in the supply and consequent damage to the equipment. A fuse will protect when excess current is drawn, but might be too slow to cope with overvoltage conditions. The values shown are for a 12 V supply being dropped to 5 V. The trip voltage is set to 5.7 V to protect the equipment in the event of a regulator fault. The 330 ohm resistor and the 500 ohm potentiometer form a potential divider which samples the output voltage as set by adjustment of the potentiometer. The SCR is selected to carry at least twice the fuse rating. The full supply voltage is connected to the input of the regulator. The 2N2906 is held bias off by the 10 k resistor and the SCR so that the LED is held off. If the output voltage rises above a set trip value then the SCR will conduct, the fuse will blow, and the 2N3906 will be supplied with base current via the 10 k resistor, and the LED will light up.

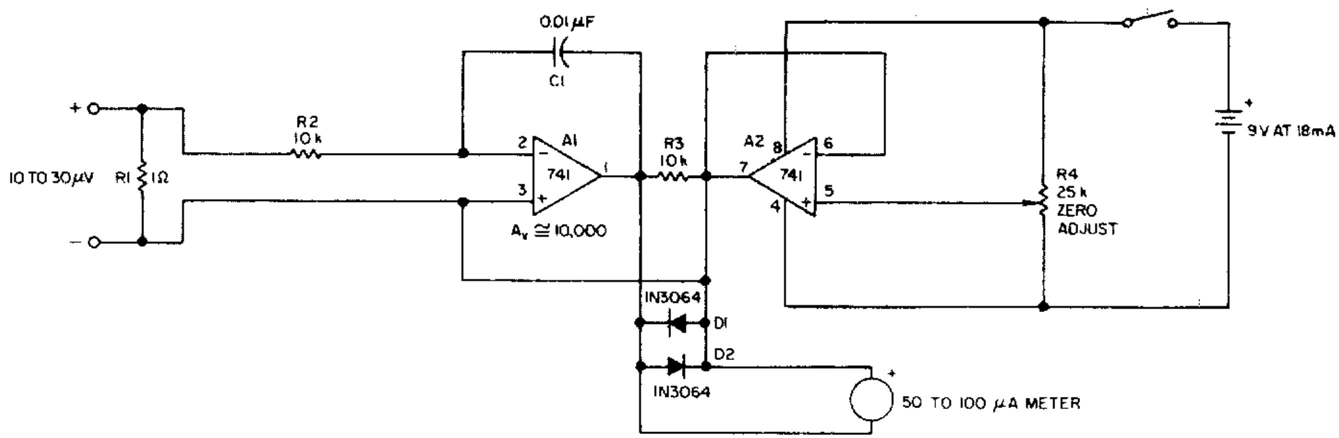
75

Probes

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Microvolt Probe
Single Injector-Tracer
General Purpose RF Detector
Clamp-on-Current Probe Compensator
650 MHz Amplifying Prescaler Probe
Tone Probe for Testing Digital ICs

MICROVOLT PROBE



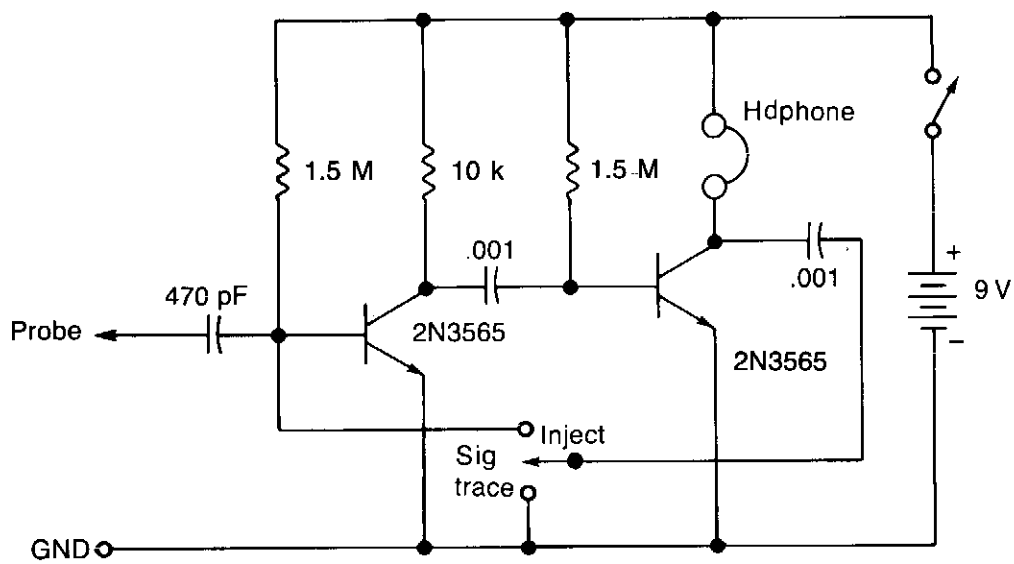
ELECTRONIC DESIGN

Fig. 75-1

Circuit Notes

The current tracer helps locate a defective IC that is loading down the power supply. The tracer amplifies the small voltage drop caused by current flow along a fraction of an inch of PC wiring and drives an ordinary microammeter. Needle-point test probes are used to contact the edge of a PC trace and to follow the current to determine which branch the current takes. One-half of a dual 741 op amp forms a dc amplifier with ac feedback to prevent oscillations and hum-pickup problems. It drives a 50-to-100 μA meter. The other op amp provides a center tap for the 9 V battery supply and zero adjustment with R4. Two diodes protect the meter. Resistor R1 eliminates the necessity for shorting the probes when the meter is zeroed. The value of 1 ohm is large when compared with the resistance of the meter leads plus the bridged portion of PC wiring.

SINGLE INJECTOR-TRACER



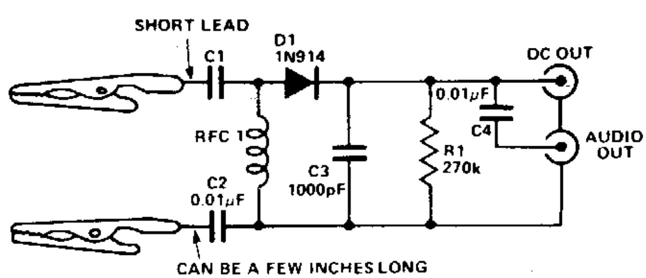
WILLIAM SHEETS

Fig. 75-2

Circuit Notes

This circuit will provide a nominal square wave output in the audio range in the "Inject" mode, the harmonics of which should be heard at several MHz. In the "Trace" mode the non-linear operation of the amplifier will detect modulated rf signals which will be filtered by the .001 μF capacitor and heard in the headphones.

GENERAL PURPOSE RF DETECTOR



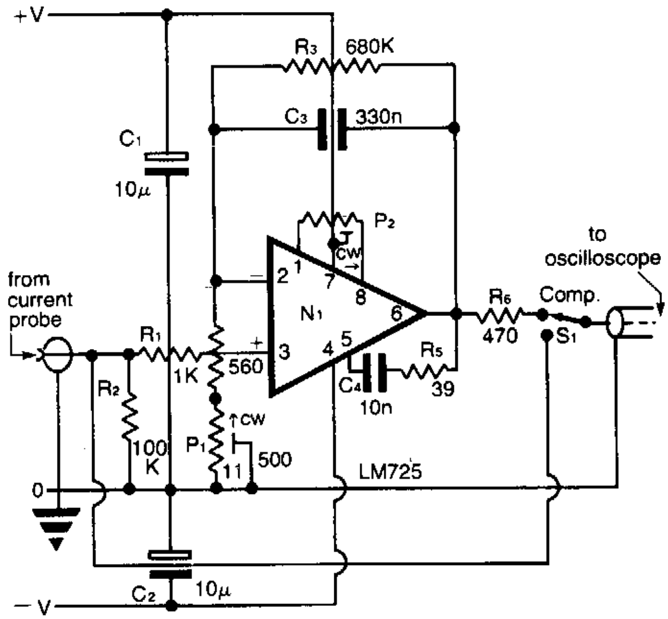
ELECTRONICS TODAY INTERNATIONAL

Fig. 75-3

Circuit Notes

This circuit provides a dc output to a meter and an audio output (if necessary) for checking transmitters or modulated signals. It can be used also as a field strength meter or transmitter monitor.

CLAMP-ON-CURRENT PROBE COMPENSATOR



Table

Tek P6021, on its own	with Tek amp. 134	with compensator
120Hz to 60MHz @ 10mA/mV 450Hz to 60MHz @ 2mA/mV	12Hz to 38MHz (switched 1mA to 1A/div for 50mV/div output	1Hz to 100kHz @ 2mA/mV

ELECTRONIC ENGINEERING

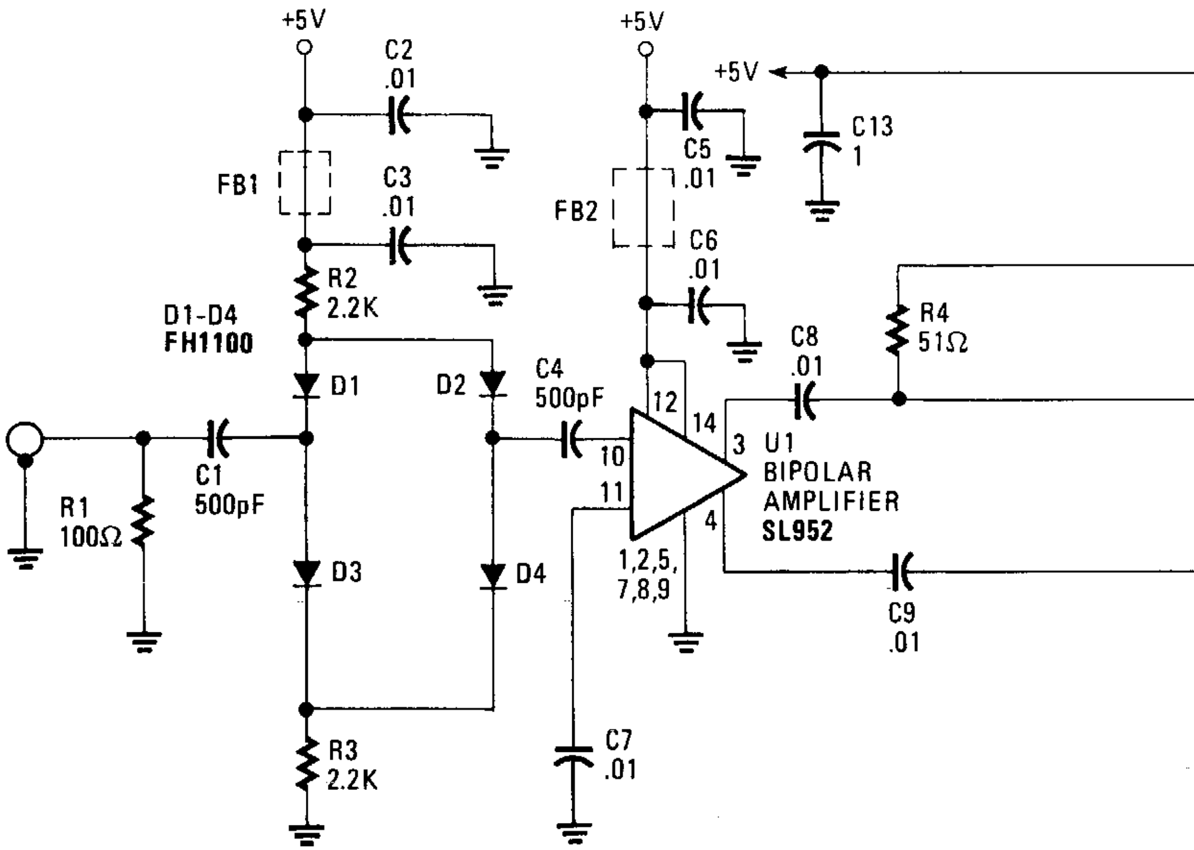
Fig. 75-4

Circuit Notes

A clamp-on "current probe" such as the Tektronix P6021 is a useful means of displaying current waveforms on an oscilloscope. Unfortunately, the low-frequency response is somewhat limited, as shown in the Table.

The more sensitive range on the P6021 is 2 mA/mV, but it has a roll-off of 6 dB per octave below 450 Hz. The compensator counteracts the low-frequency attenuation, and this is achieved by means of C3 and R4 + P1 in the feedback around op amp N1. The latter is a low-noise type, such as the LM725 shown, and even so it is necessary at some point to limit the increasing gain with decreasing frequency; otherwise amplifier noise and drift will overcome the signal. The values shown for C3R3 give a lower limit below 1 Hz. A test square wave of ±10 mA is fed to the current probe so that P1 can be adjusted for minimum droop or overshoot in the output waveform. At high frequencies, the response begins to fall off at 100 kHz.

650 MHz AMPLIFYING PRESCALER PROBE

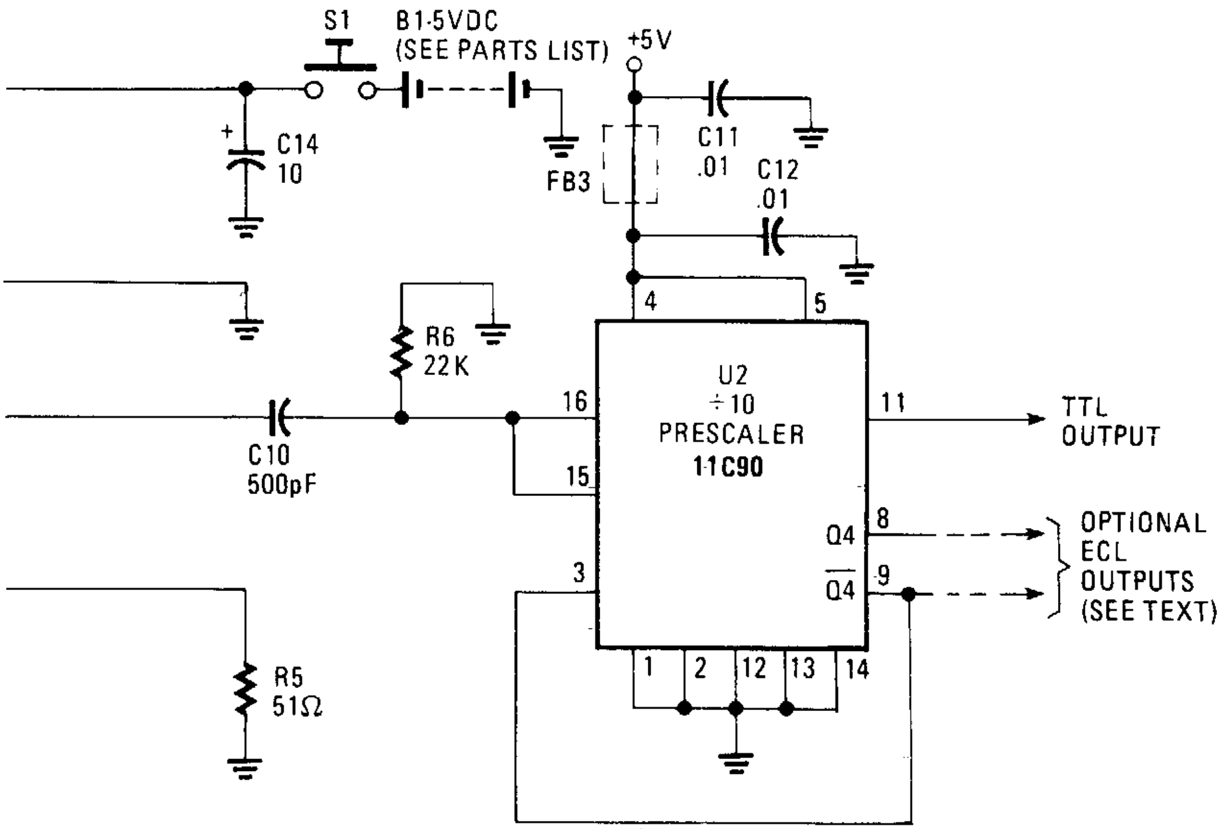


HANDS-ON ELECTRONICS

Fig. 75-5

Circuit Notes

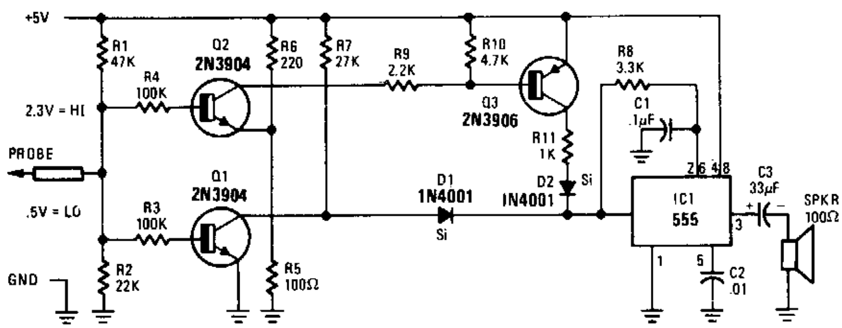
The 650 MHz Prescaler Probe's input is terminated by resistor R1 and is fed through C1 to the diode limiter composed of D1 through D4. Those diodes are forward-biased by the +5 volt supply for small-input signals and, in turn, feed the signal to U1. However, for larger input signals, diodes D1 through D4 will start to turn off, passing less of the signal, and, thus, attenuating it. But even in a full-off state, the FH1100-type diodes will always pass a small part of the input to U1 because of capacitive leakage within the diodes. Integrated circuit U1, a Plessey SL952 bipolar amplifier, capable of 1 GHz operation, provides 20 to 30 dB of gain. The input signal is supplied to pin 10, U1 with



the other input (pin 11) is bypassed to ground. The output signal is taken at pin 3 and pin 4, with pin 3 loaded by R4 and pin 4 by R5.

Integrated circuit 11C90, U2, is a high-speed prescaler capable of 650 MHz operation configured for a divide-by-10 format. A reference voltage internally generated appears at pin 15 and is tied to pin 16, the clock input. This centers the capacitive-coupled input voltage from U1 around the switching threshold-voltage level. An ECL-to-TTL converter in U1 provides level conversion to drive TTL input counters by typing pin 13 low. Therefore, no external ECL to TTL converter is required at the pin 11 output. On the other hand, ECL outputs are available at U2, pin 8 (Q4) and at pin 9 (Q4), if desired. In that circuit configuration, pin 13 is left open, and U2 will use less power.

TONE PROBE FOR TESTING DIGITAL ICs



RADIO-ELECTRONICS

Fig. 75-6

Circuit Notes

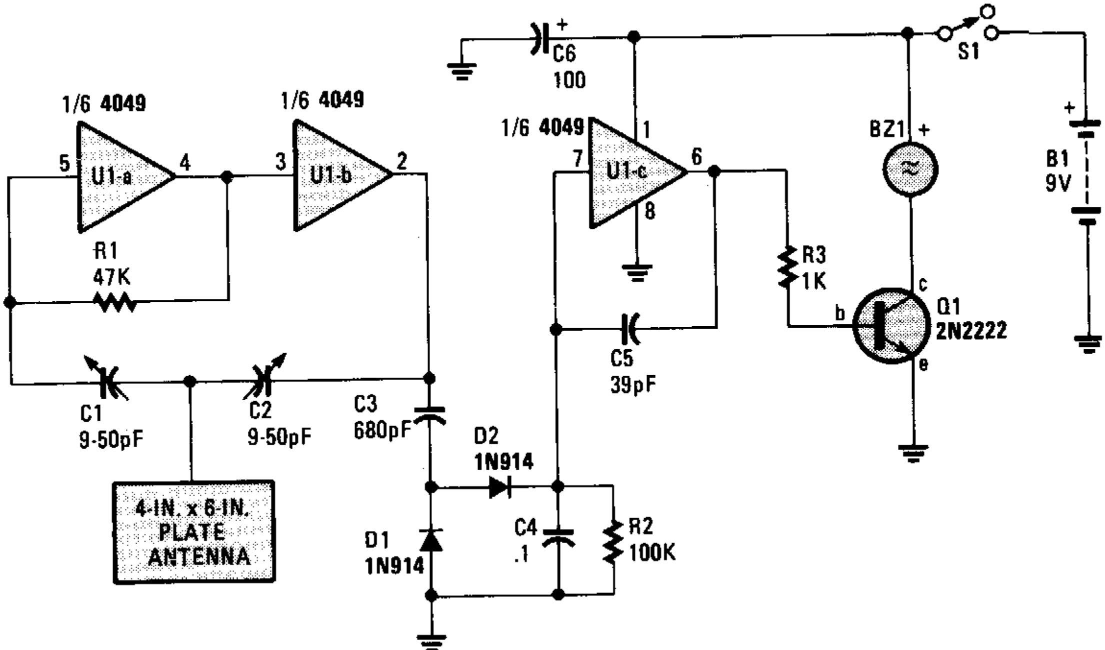
The tone probe uses sound to tell the status of the signal being probed. The probe's input circuit senses the condition of the signal and produces either a low-pitched tone for low-level signals (less than 0.8 V) or a high-pitched tone for high-level signals (greater than 2 V).

76

Proximity Sensors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

PROXIMITY ALARM



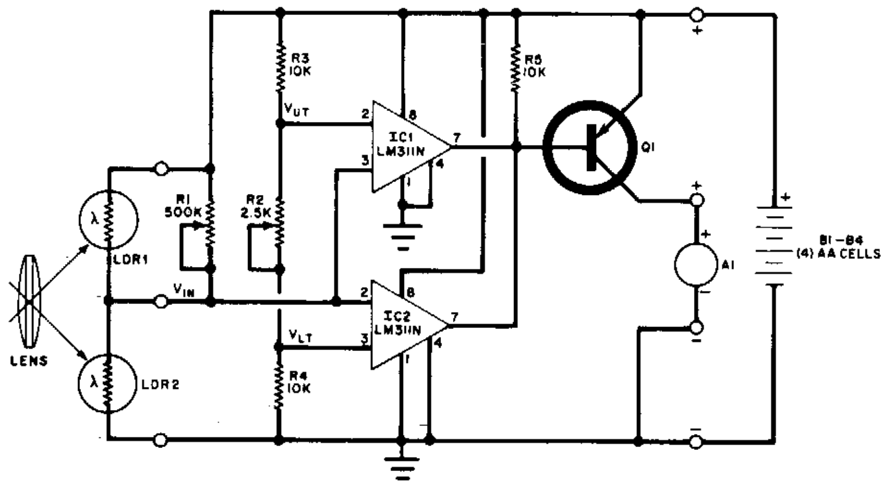
HANDS-ON ELECTRONICS

Fig. 76-1

Circuit Notes

Inverters U1a and U1b are connected in a simple RC oscillator circuit. The frequency is determined by the values of R1, C1, C2, and the internal characteristics of the integrated circuit. As long as the circuit is oscillating, a positive dc voltage is developed at the output of the voltage-coupler circuit: C3, D1, D2 and C4. The dc voltage is applied to the input of U1c—the third inverter amplifier—keeping its output in a low state, which keeps Q1 turned off so that no sound is produced by BZ1. With C1 and C2 adjusted to the most sensitive point, the pickup plate will detect a hand 3 to 5-inches away and sound an alert. Set C1 and C2 to approximately one-half of their maximum value and apply power to the circuit. The circuit should oscillate and no sound should be heard. Using a non-metallic screwdriver, carefully adjust C1 and C2, one at a time, to a lower value until the circuit just ceases oscillation: Buzzer BZ1 should sound off. Back off either C1 or C2 just a smidgen until the oscillator starts up again—that is the most sensitive setting of the circuit.

FIELD DISTURBANCE SENSOR/ALARM



POPULAR ELECTRONICS *Fig. 76-2*

Circuit Notes

The change in ambient light triggers the alarm by changing resistance of LDR1 and LDR2.

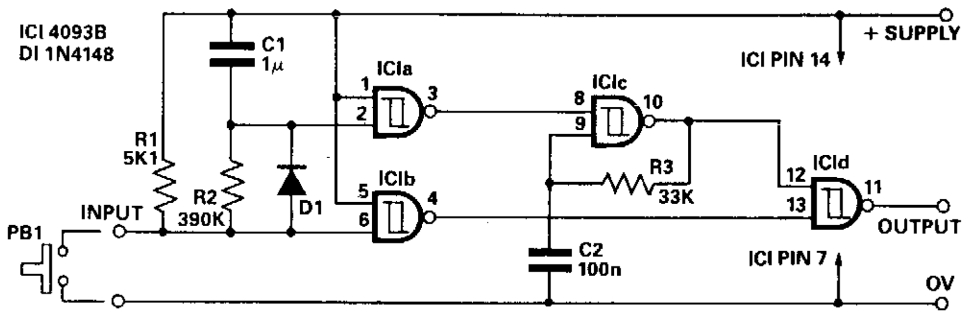
Q1 = Radio Shack 276-2024
A1 = Mallory SC628P Sonalert
LDR1, LDR2 = Cadmium sulfide photocell, Radio Shack 276-116

77

Pulse Generators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DELAYED PULSE GENERATOR



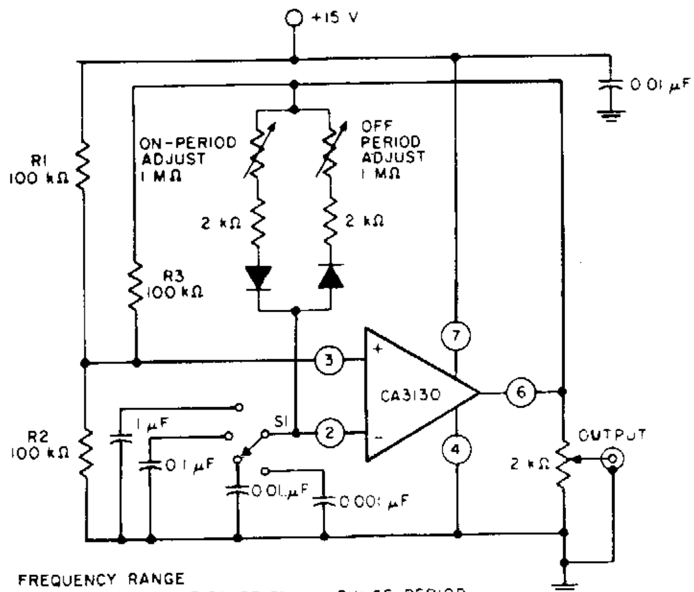
ELECTRONIC ENGINEERING

Fig. 77-1

Circuit Notes

The circuit offers independent control of initial delay and pulse rate. IC1c is connected as a pulse generator whose operation is inhibited by the normally low O/P of the IC1a. When the circuit input goes low i.e., by pressing PB1, IC1b O/P goes high and the circuit O/P goes low thus replicating the input. When the input is kept low capacitor C1 charges via R2 to a point where IC1a O/P goes low. This allows the pulse generator IC1c to start and "rapid fire" pulses appear at the circuit O/P. When the circuit input returns to the high state C1 is rapidly discharged via D1 and R1. The value of R2 and C1 control the initial delay while R3 and C2 control the pulse rate. The values given will give a delay of around 0.5 seconds and a pulse rate of 200/300 Hz depending on supply voltage. PB1 may be replaced by an open collector TTL gate or a common emitter transistor stage if required.

PULSE GENERATOR (ASTABLE MULTIVIBRATOR)



FREQUENCY RANGE

POSITION OF S1	PULSE PERIOD
0.001 µF	4 µs TO 1 ms
0.01 µF	40 µs TO 10 ms
0.1 µF	0.4 ms TO 100 ms
1 µF	4 ms TO 1 s

RCA

Fig. 77-2

Circuit Notes

Resistors R1 and R2 bias the CA3130 to the mid-point of the supply-voltage, and R3 is the feedback resistor. The pulse repetition rate is selected by positioning S1 to the desired position and the rate remains essentially constant when the resistors which determine on-period and off-period are adjusted.

NON-INTEGER PROGRAMMABLE PULSE DIVIDER

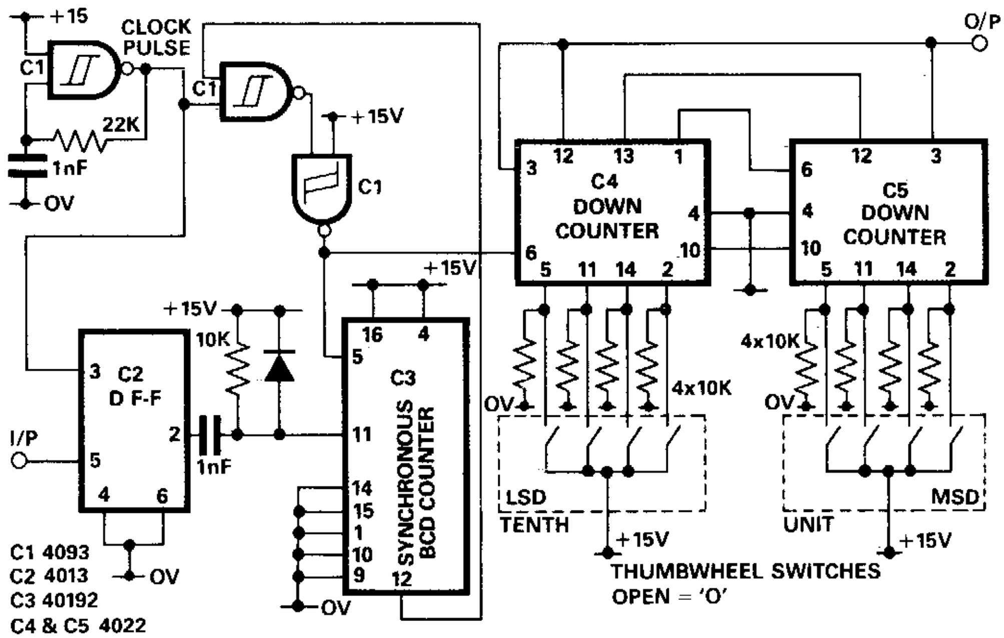


Fig. 77-3

ELECTRONIC ENGINEERING

Circuit Notes

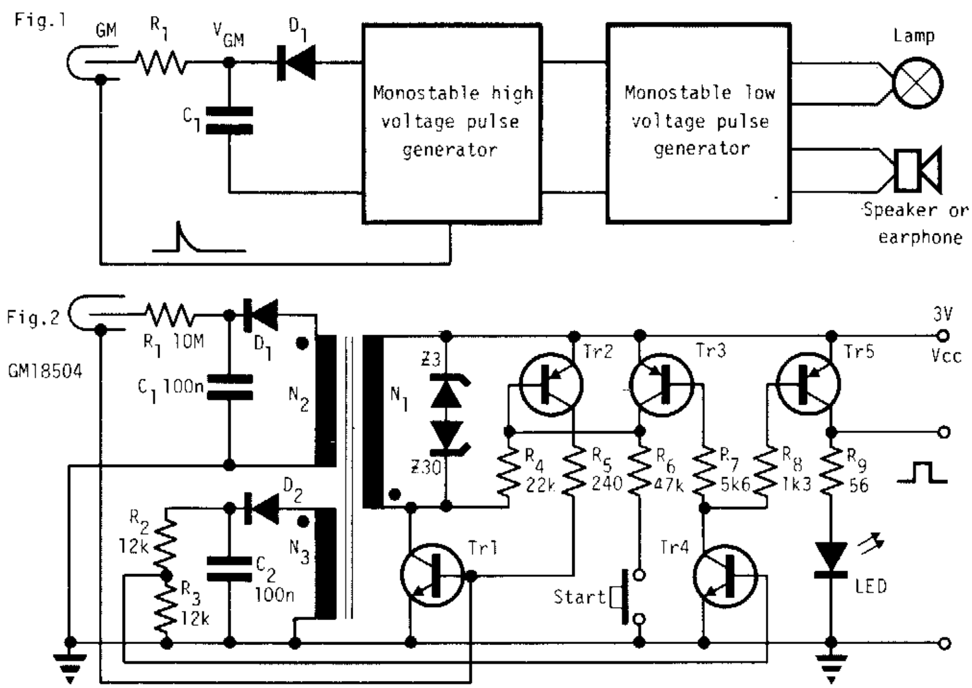
In applications where the period of the input pulses is uneven and the divider is required to cover a wide range of frequencies, the non-integer programmable pulse divider shown can be used. The purpose of the D-type flip-flop (IC2) is to synchronize the input signal with the clock pulse. When the clock pulse changes from low to high and the input is high, IC2 output goes high. Subsequently, IC3 resets to zero and starts counting up. The number of pulses at the output of IC3 is ten time the input pulse. IC4 and IC5 are cascaded to form a two decade programmable down counter.

78

Radiation Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

MICROPOWER RADIOACTIVE RADIATION DETECTOR



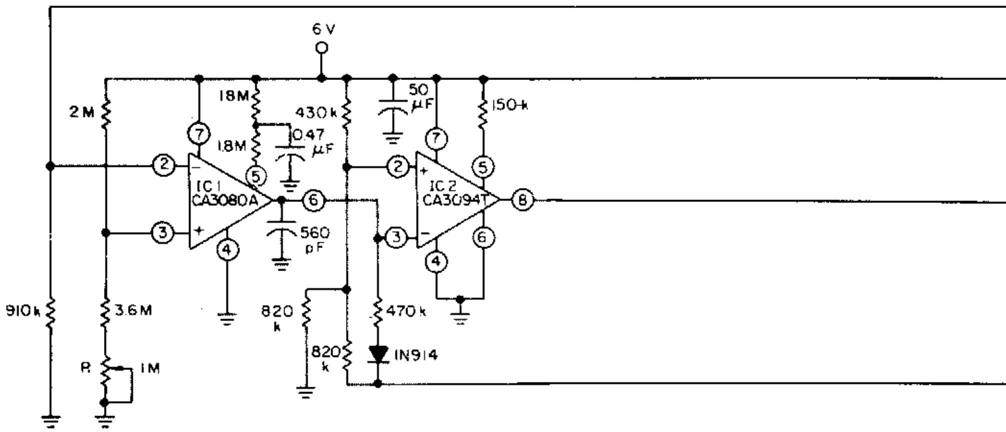
ELECTRONIC ENGINEERING

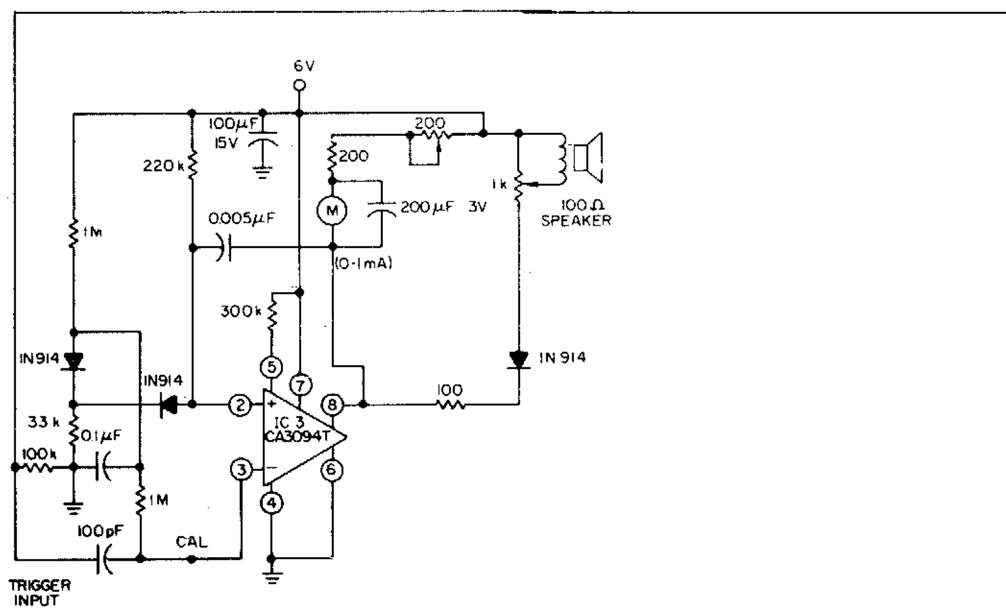
Fig. 78-1

Circuit Notes

In the absence of radiation, no current is drawn. At normal background radiation levels the power consumption is extremely low. The instrument may be left on for several months without changing batteries. In this way the detector is always ready to indicate an increase in radiation. An LED is used as an indicator lamp. With background radiation it draws less than 50 μA. A ferrite pot core is used for the transformer with N1 = 30, N2 = 550, and N3 = 7. Using two 1.5V batteries with 0.5 Ah total capacity, the detector can work at background radiation levels for 0.5 Ah $\div$ 50μA = 10,000 hours, which is more than a year.

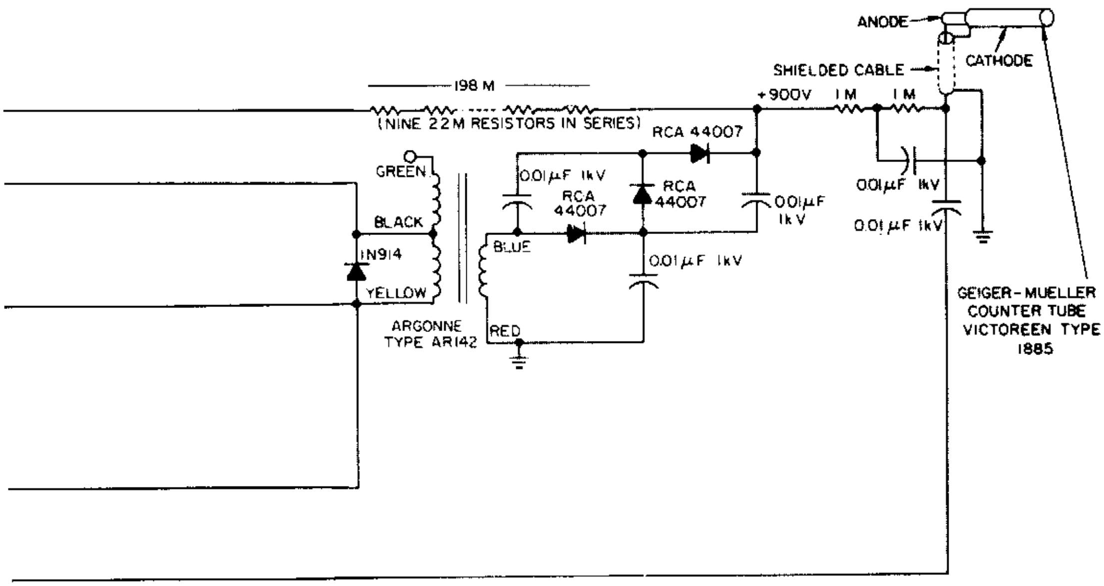
POCKET-SIZED GEIGER COUNTER





ELECTRONIC DESIGN

Fig. 78-2



Circuit Notes

A single 6.75 V mercury battery powers the counter, which features a 1 mA count-rate meter as well as an aural output. A regulated 900 V supply provides stable operation of the counter tube. A multivibrator, built around a differential power amplifier IC2, drives the step-up transformer. Comparator IC1 varies the multivibrator duty cycle to provide a constant 900 V. The entire regulated supply draws less than 2 mA. A one-shot multivibrator, built with IC3, provides output pulses that have constant width and amplitude. Thus the average current through the meter is directly proportional to the pulse-rate output from the counter tube. And the constant-width pulses also drive the speaker. Full-scale meter deflection (1 mA) represents 5000 counts/min, or 83.3 pulses/s. A convenient calibration checkpoint can be provided on the meter scale for 3600 ppm (60 pulses/s.)

PHOTOMULTIPLIER OUTPUT-GATING CIRCUIT

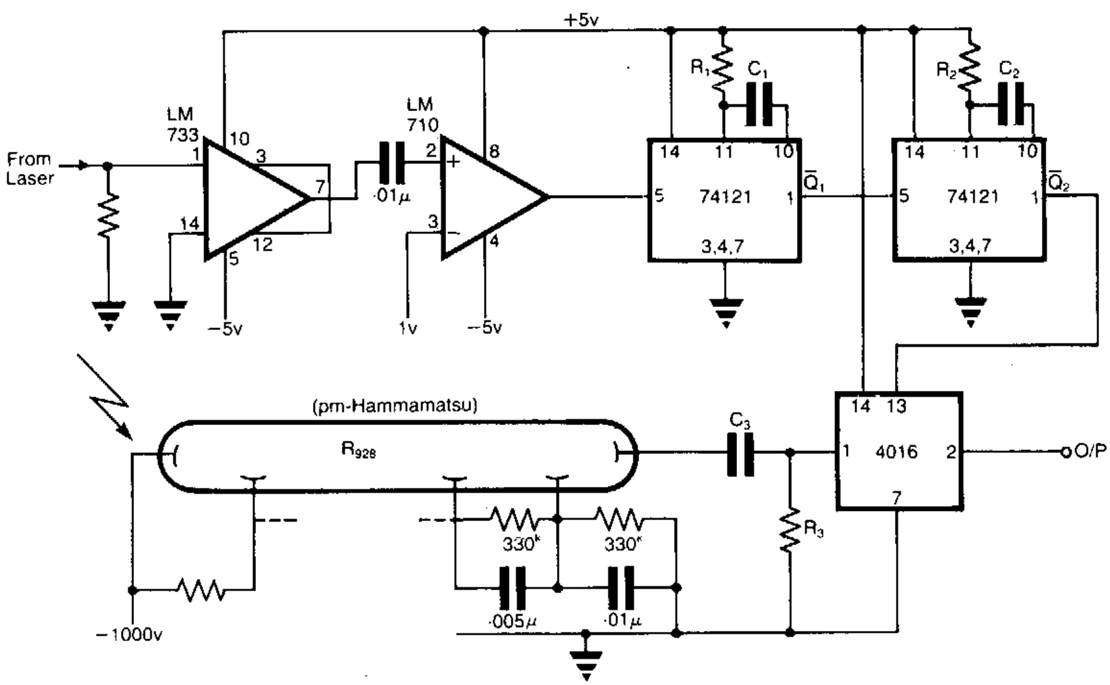


Fig. 78-3

ELECTRONIC ENGINEERING

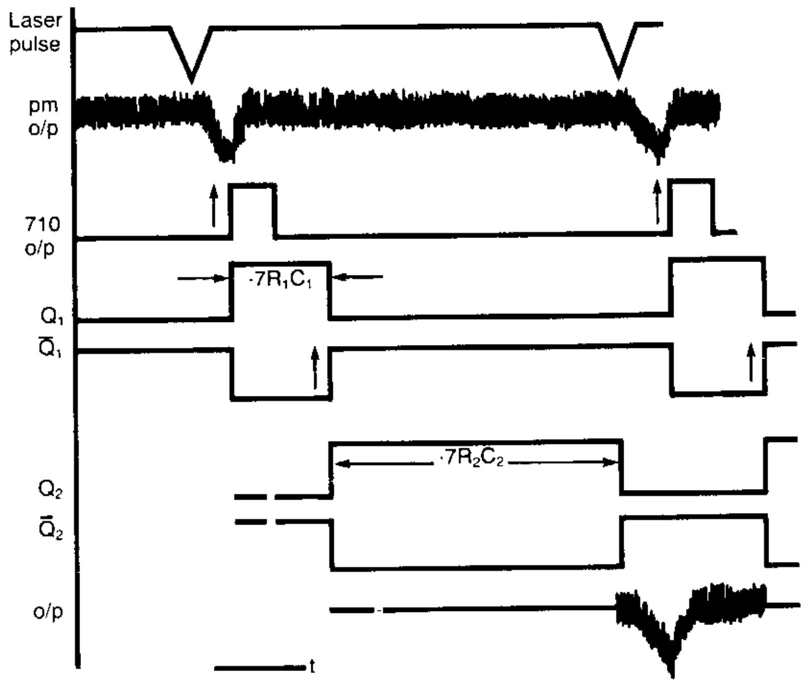


Fig 2

Circuit Notes

The application involves observing the light pulse emerging from a thick specimen after transillumination by a laser pulse. Pulses derived from the laser source are amplified using a Video Amplifier LM733. The reference level is set to 1 V in the comparator LM 710, to provide the necessary trigger pulses for the monostable multivibrator 74121. The laser pulses have a repetition frequency of 500 Hz and suitable values are as below:

R1 = 33 k ohm, C1 = 22 pF
R2 = 33 k ohm, C2 = 68 nF

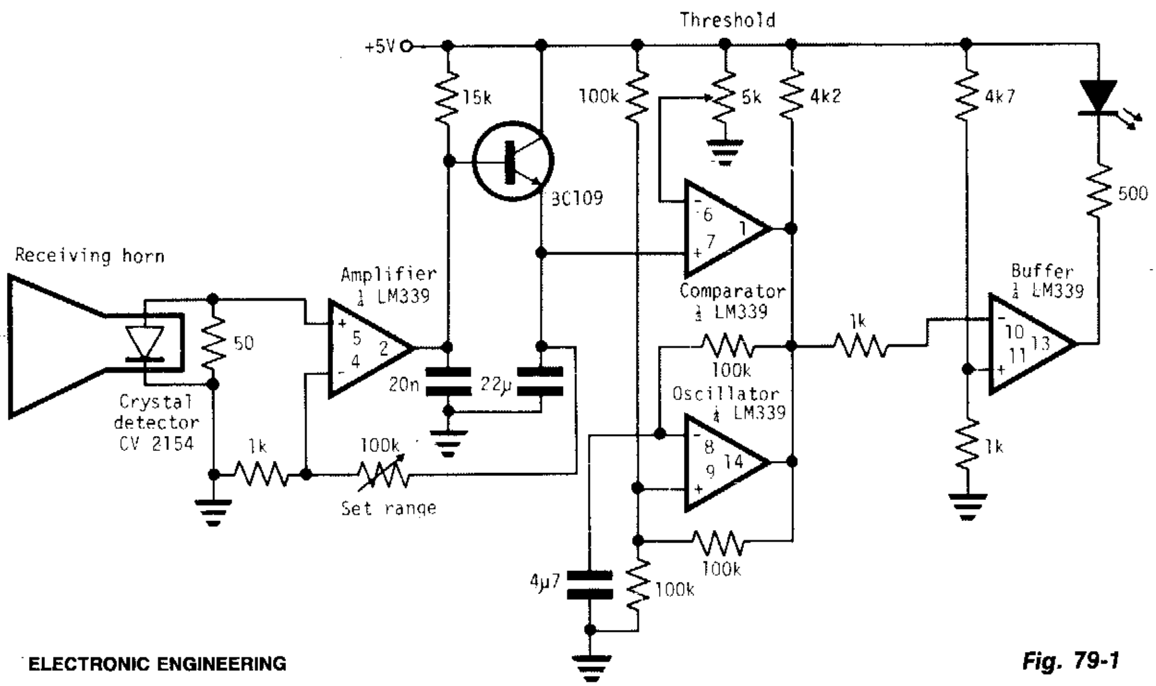
The pulse width for each monostable is approximately given by tw = 0.7 RC. R3 and C3 is a high pass filter. The method therefore permits the use of low cost components having moderate response times for extracting the pulse of interest.

79

Radar Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ONE-CHIP RADAR DETECTION CIRCUIT



ELECTRONIC ENGINEERING

Fig. 79-1

Circuit Notes

A simple X-band radar detector is capable of indicating changes in rf radiation strength at levels down to 2 mW/cm^2. Radiation falling on the detector diode, produces a voltage at the input of an amplifier whose gain may be adjusted to vary the range at which the warning is given. The amplifier output drives a voltage comparator with a variable threshold set to a level that avoids false alarms. The comparator output is connected in the wired-OR configuration with the open collector output of an oscillator running at a frequency of 2 Hz. In the absence of a signal, the comparator output level is low, inhibiting the oscillator output stage and holding the buffer so the lamp is off. When a signal appears, the comparator output goes high, removing the lock from the oscillator which free-runs, switching the lamp on and off at 2 Hz.

RADAR SIGNAL DETECTOR

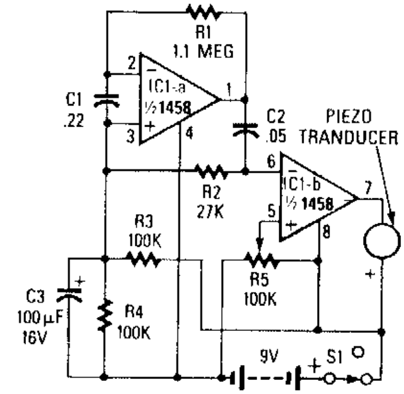


FIG. 1—THE ECONOMY RADAR DETECTOR needs only one IC and a few discrete components.

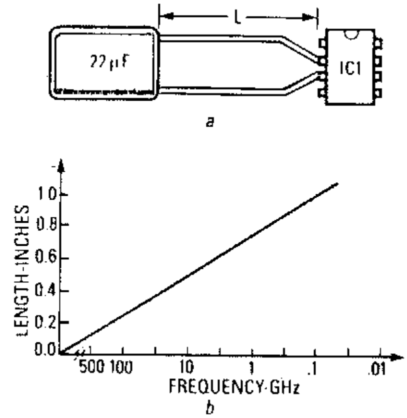


FIG. 3—VARY THE LEAD LENGTHS OF C1 to tune the input circuit.

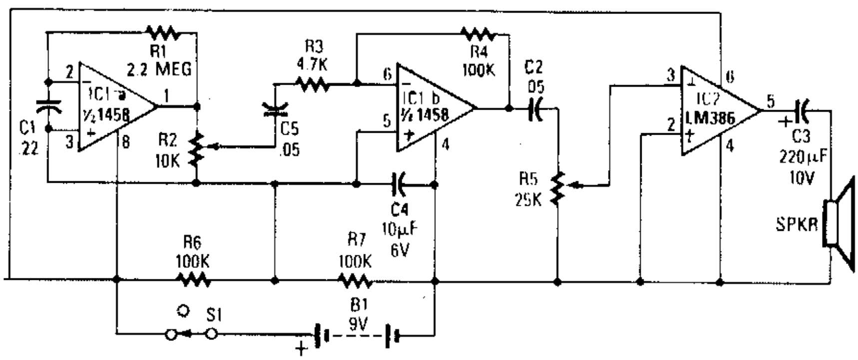


FIG. 2—DELUXE RADAR DETECTOR adds a buffer amplifier and an audio power amp to drive a speaker.

RADIO-ELECTRONICS

Fig. 79-2

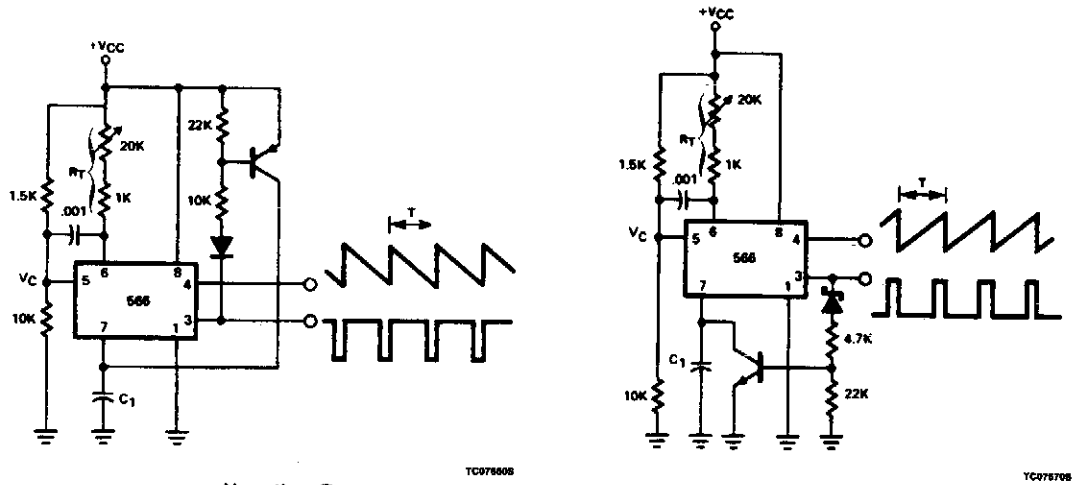
Circuit Notes

The circuit can be tuned to respond to signals between 50 MHz and 500 GHz. The economy model is shown in Fig. 1, and the deluxe model is shown in Fig. 2. The first op amp in each circuit functions as a current-to-voltage converter. In the economy model IC1b buffers the output to drive the piezo buzzer. The deluxe model functions in a similar manner except that IC1b is configured as a ×20 buffer amplifier to drive the LM386. In both circuits C1 functions as a "transmission line" that intercepts the incident radar signal. The response may be optimized by trimming C1's lead length for the desired frequency. Typically the capacitor's leads should be 0.5-0.6 inches long.

80
Ramp Generators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RAMP GENERATOR



a. Negative Ramp

b. Positive Ramp

SIGNETICS

Fig. 80-1

Circuit Notes

The 566 can be wired as a positive or negative ramp generator. In the positive ramp generator, the external transistor driven by the Pin 3 output rapidly discharges C1 at the end of the charging period so that charging can resume instantaneously. The pnp transistor of the negative ramp generator likewise rapidly charges the timing capacitor C1 at the end of the discharge period. Because the circuits are reset so quickly, the temperature stability of the ramp generator is excellent. The period

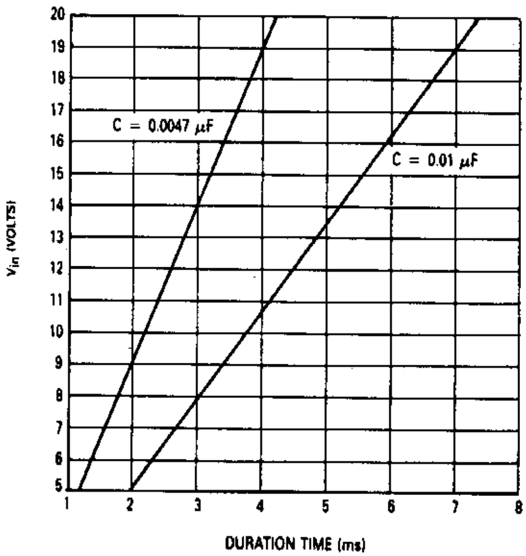
$$T \text{ is } \frac{1}{2f_o}$$

where f_o is the 566 free-running frequency in normal operation. Therefore,

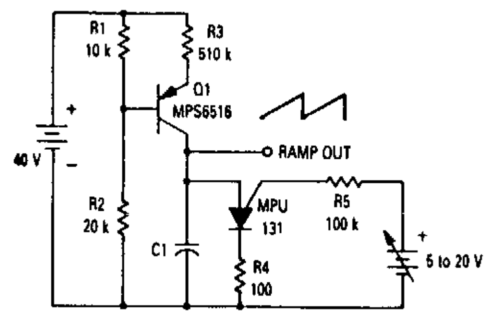
$$T = \frac{1}{2f_o} = \frac{R_T C_1 V_{CC}}{.5(V_{CC} - V_C)} \quad (1)$$

where V_C is the bias voltage at Pin 5 and R_T is the total resistance between Pin 6 and V_{CC}. Note that a short pulse is available at Pin 3. (Placing collector resistance in series with the external transistor collector will lengthen the pulse.)

VOLTAGE CONTROLLED RAMP GENERATOR



Voltage versus Ramp Duration Time of VCRG



Voltage Controlled Ramp Generator (VCRG)

MOTOROLA

Fig. 80-2

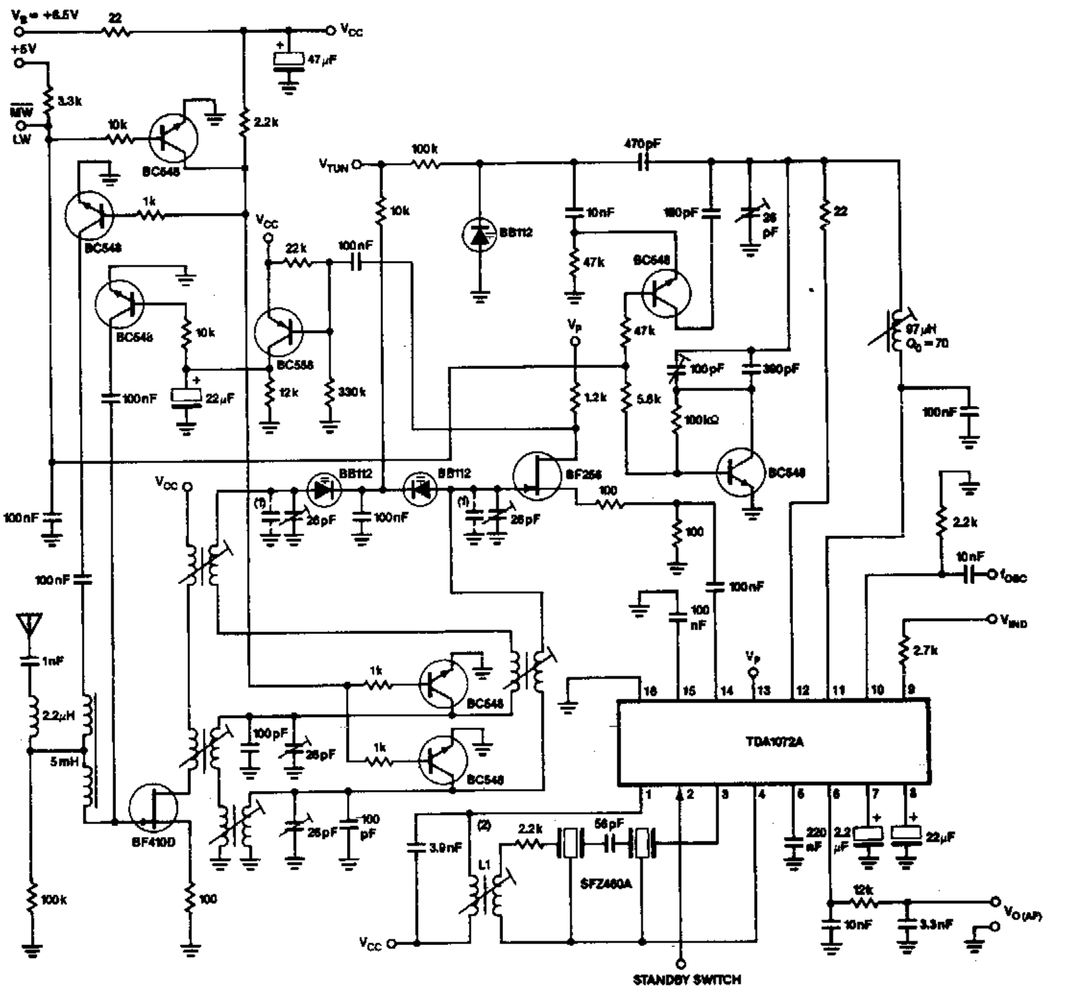
Circuit Notes

The current source formed by Q1 in conjunction with capacitor C1 set the duration time of the ramp. As the positive dc voltage at the gate is changed, the peak point firing voltage of the PUT is changed, which changes the duration time, i.e., increasing the supply voltage increases the peak point firing voltage causing the duration time to increase.

81
Receivers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

CAR RADIO WITH CAPACITIVE DIODE TUNING AND ELECTRONIC MW/LW SWITCHING



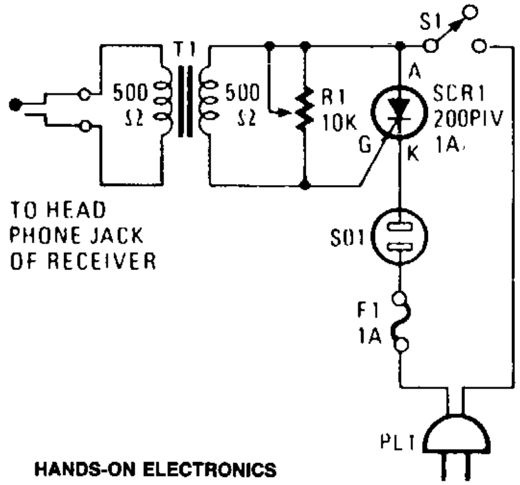
TC12991S

NOTES:
1. Values of capacitors depend on the selected group of capacitive diodes BB112.
2. For IF filter and coil data refer to Block Diagram.
3. The circuit includes pre-stage AGC optimized for good large-signal handling.

SIGNETICS

Fig. 81-1

RECEIVER MONITOR



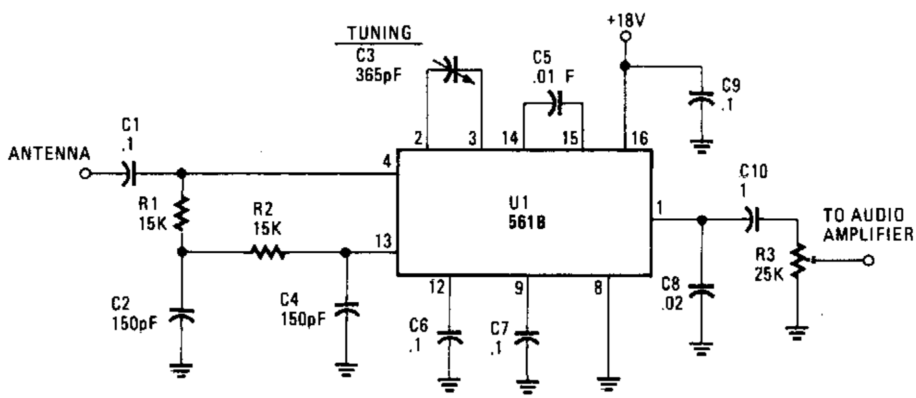
HANDS-ON ELECTRONICS

Fig. 81-2

Circuit Notes

The alarm plugs into the earphone jack on a receiver. Then when a signal (normally fed to the headphones) is detected and applied to the gate of SCR1, it conducts, sounding whatever alarm is connected to SO1. The signaling device can be an audible alarm or a lamp. Variable resistor R1 functions as a sensitivity control so that background noises won't trigger the alarm.

PLL/BC RECEIVER



HANDS-ON ELECTRONICS

Fig. 81-3

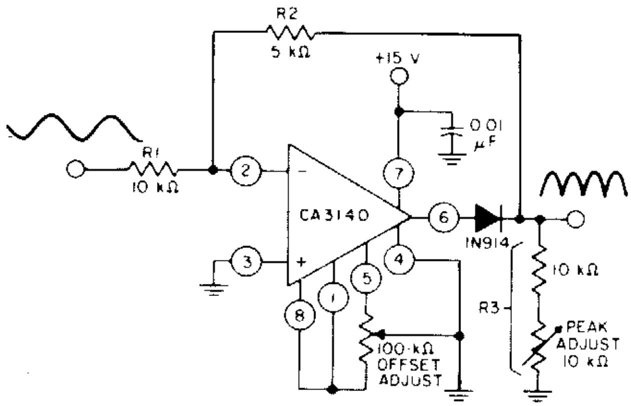
Circuit Notes

This simple AM circuit uses a 561B. There's no inductance/capacitance tuning circuit. The 365 pF capacitor connected between pins 2 and 3 does all the tuning. The circuit needs a good outside antenna and a solid ground. And if you want to further improve operation, stick a broadband amplifier in front of the receiver. Just make sure the input voltage does not climb over 0.5 volt rms.

82
Rectifier Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ABSOLUTE-VALUE, "IDEAL" FULL-WAVE RECTIFIER



$$GAIN = \frac{R2}{R1} = X = \frac{R3}{R1 + R2 + R3}$$

$$R3 = R1\left(\frac{X + X^2}{1 - X}\right)$$

$$FOR\ X = 0.5\quad \frac{5\,k\Omega}{10\,k\Omega} = \frac{R2}{R1}$$

$$R3 = 10\,k\Omega\left(\frac{0.75}{0.5}\right) = 15\,k\Omega$$

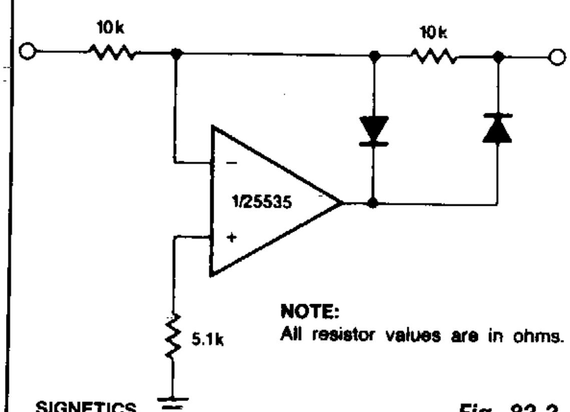
20 V p-p INPUT BW(−3dB) = 290 kHz, DC OUTPUT (AVG) = 3.2 V

GENERAL ELECTRIC/RCA ***Fig. 82-1***

Circuit Notes

The circuit uses a CA3140 BiMOS op amp in an inverting gain configuration. When equality of two equations shown in satisfied, full-wave output of circuit is symmetrical.

HALF-WAVE RECTIFIER



NOTE:
All resistor values are in ohms.

SIGNETICS ***Fig. 82-2***

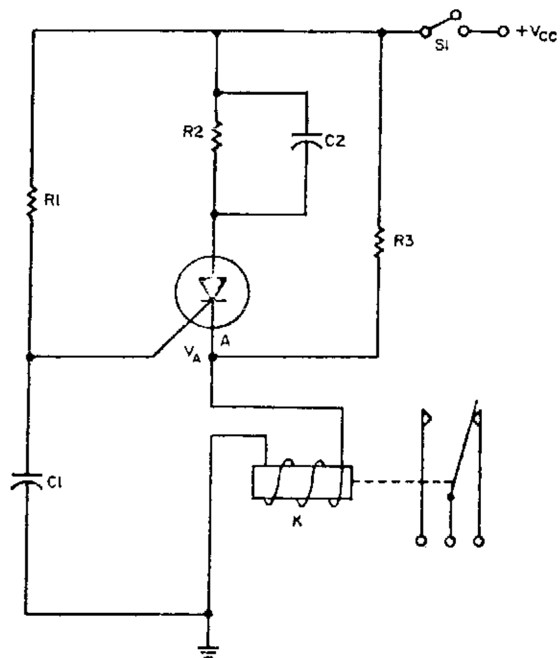
Circuit Notes

The circuit provides for accurate half-wave rectification of the incoming signal. For positive signals, the gain is 0; for negative signals, the gain is −1. By reversing both diodes, the polarity can be inverted. This circuit provides an accurate output, but the output impedance differs for the two input polarities and buffering may be needed. The output must slew through two diode drops when the input polarity reverses. The NE5535 device will work up to 10 kHz with less than 5% distortion.

83
Relay Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

RELAY DRIVER PROVIDES DELAY AND CONTROLS CLOSURE TIME



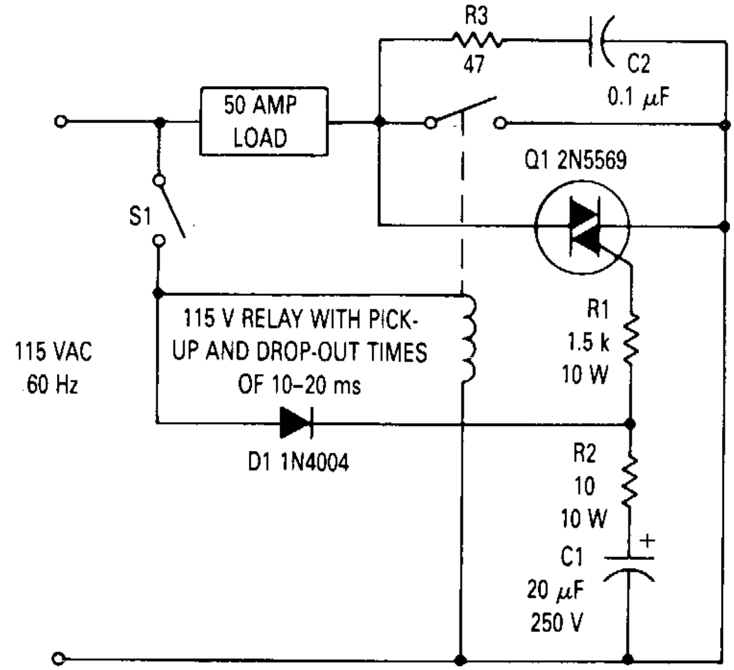
ELECTRONIC DESIGN

Fig. 83-1

Circuit Notes

The relay operates a certain time, t_d, after power is applied to it, and then it operates for a length of time, t_c. The SCR fires when the voltage on C1 reaches V_A. This operates the relay, which stays activated until the current charging C2 drops below the dropout current. To keep the relay in its activated position indefinitely ($t_c = \infty$), eliminate C2 and choose R2 just large enough to keep the relay coil current within its related limits. Typical component values for t_d = 30 seconds and t_c = 2 seconds are: R1 = 1.5 megohms, R2 = 10 k ohms, R3 = 3 k ohms, C1 = 47 μF, and C2 = 100 μF. The SCR is a 2N1877 and the relay is a Potter Brumfield PW-5374. A value of 12 Vdc is assumed for V_{cc}.

TRIAC RELAY-CONTACT PROTECTION



MOTOROLA

Fig. 83-2

Circuit Notes

This circuit can be used to prevent relay contact arcing for loads up to 50 amperes. There is some delay between the time a relay coil is energized and the time the contacts close. There is also a delay between the time the coil is de-energized and the time the contacts open. For the relay used in this circuit both times are about 15 ms. The TRIAC across the relay contacts will turn on as soon as sufficient gate current is present to fire it. This occurs after switch S1 is closed but before the relay contacts close. When the contacts close, the load current passes through them, rather than through the TRIAC, even though the TRIAC is receiving gate current. If S1 should be closed during the negative half cycle of the ac line, the TRIAC will not turn on immediately but will wait until the voltage begins to go positive, at which time diode D1 conducts providing gate current through R1. The maximum time that could elapse before the TRIAC turns on is 8-⅓ ms for the 60 Hz supply. This is adequate to ensure that the TRIAC will be on before the relay contact closes.

TR CIRCUIT

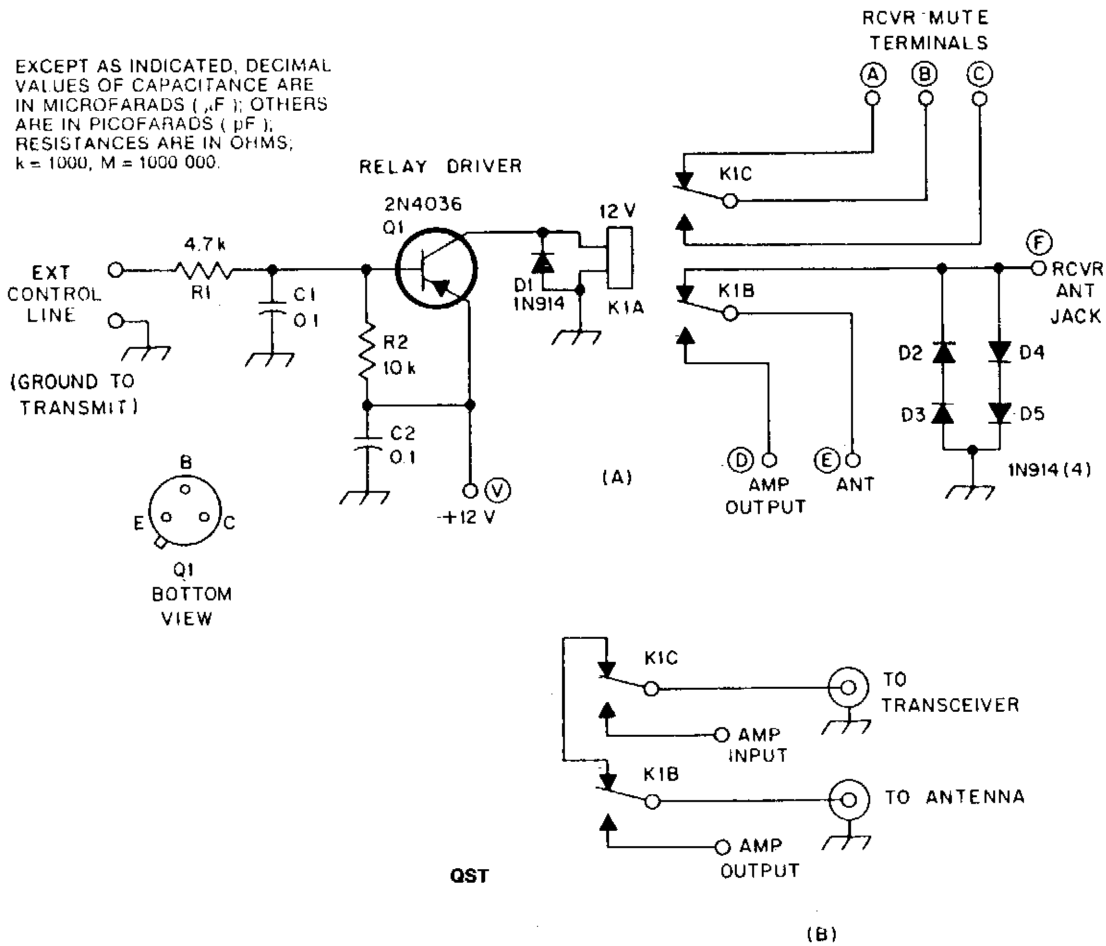


Fig. 83-3

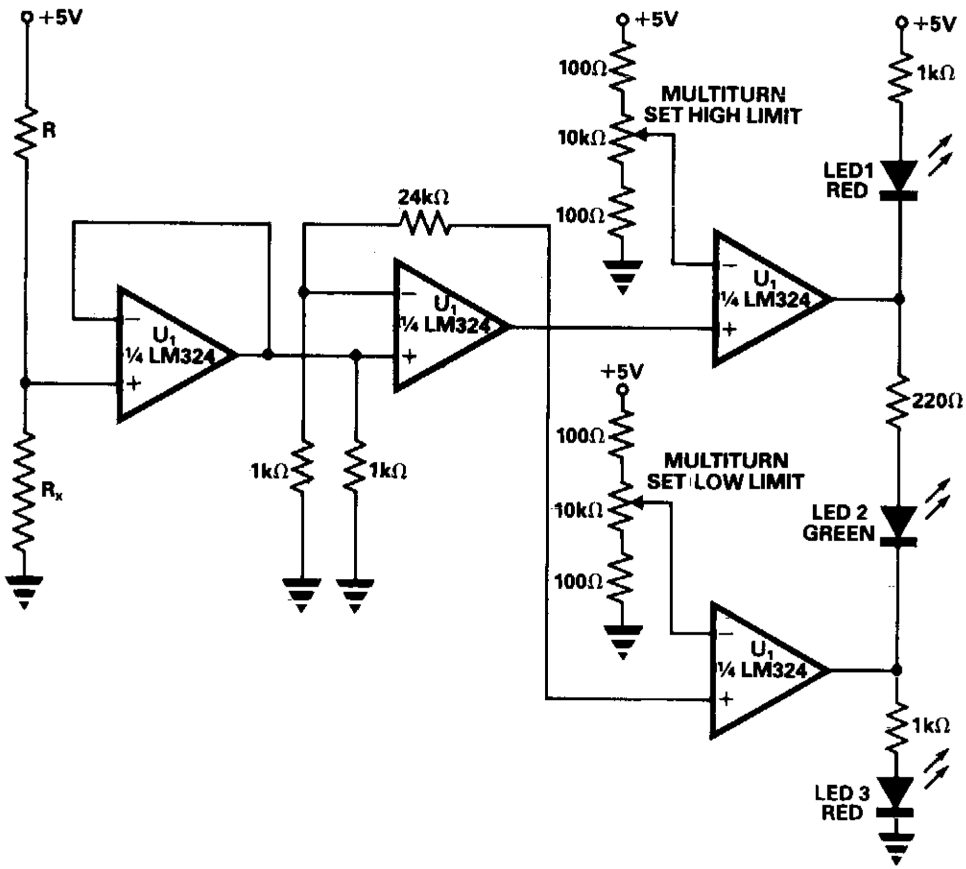
Circuit Notes

C1 and C2 are disc ceramic. R1 and R2 are ¼ or ½ W carbon composition resistors. K1 is a 12 V DPDT DIP relay. Illustration A shows how to connect the relay contacts for use with a separate transmitter-receiver combination. The circuit at B is for amplifier use with a transceiver.

84
Resistance/Continuity Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SINGLE CHIP CHECKS RESISTANCE



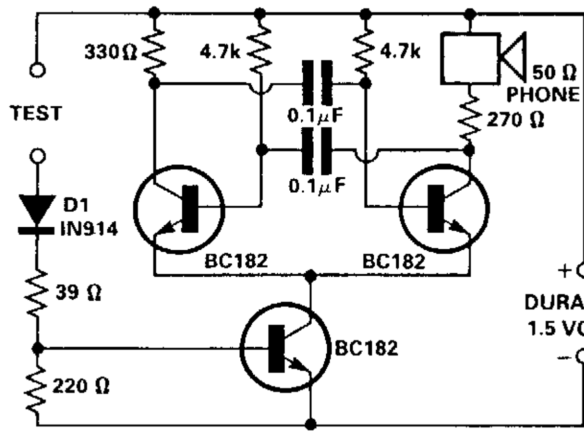
ELECTRONIC ENGINEERING

Fig. 84-1

Circuit Notes

A simple tester can be used for routine checks for resistance on production lines of relays, coils, or similar components where frequent changes in resistance to be tested are not required. The tester is built around a single quad op amp chip, the LM324. R, which is chosen to be around 80 times the resistance to be checked, and the 5 V supply form the current source. The first op amp buffers the voltage generated across the resistance under test, R_x. The second op amp amplifies this voltage. The third and fourth op amps compare the amplified voltage with high and low limits. The high and low limits are set on multiturn presets with high and low limit resistors connected in place of R_x. LED 1 (red) lights when the resistance is high. LED 2 (green) shows that the resistance is within limits. LED 3 (red) indicates that the resistance is low.

SIMPLE CONTINUITY TESTER FOR PCBs



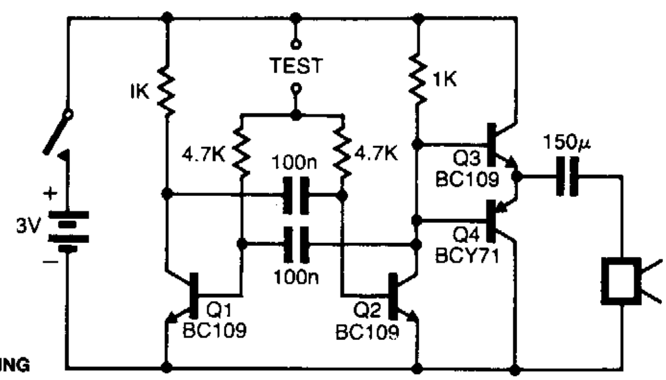
ELECTRONIC ENGINEERING

Fig. 84-2

Circuit Notes

This tester is for tracing wiring on Printed Circuit Boards. Resistors below 50 ohms act as a short circuit; above 100 ohms as an open circuit. The circuit is a simple multivibrator switched on by transistor T3. The components in the base of T3 are D1, R1, R2, and the test resistance. With a 1.5 volt supply, there is insufficient voltage to turn on a semiconductor connected to the test terminals.

SIMPLE CONTINUITY TESTER



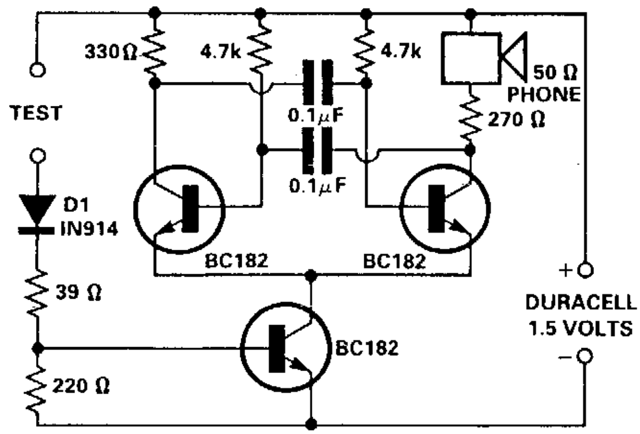
ELECTRONIC ENGINEERING

Fig. 84-3

Circuit Notes

The pitch of the tone is dependent upon the resistance under test. The tester will respond to resistance of hundreds of kilohms, yet it is possible to distinguish differences of just a few tens of ohms in low-resistance circuits. Q1 and Q2 form a multivibrator, the frequency of which is influenced by the resistance between the test points. The output stage Q3 and Q4 will drive a small loudspeaker or a telephone earpiece.

SIMPLE CONTINUITY TESTER FOR PCBs



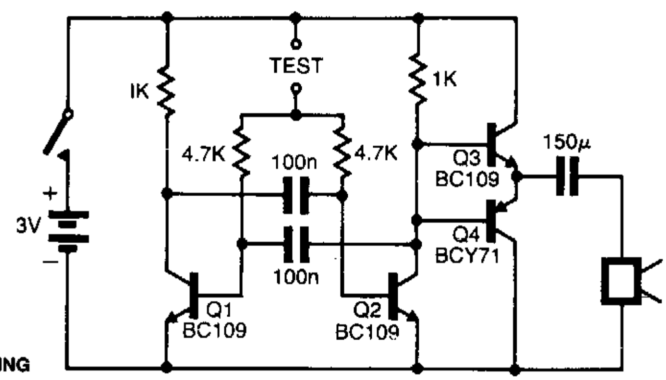
ELECTRONIC ENGINEERING

Fig. 84-2

Circuit Notes

This tester is for tracing wiring on Printed Circuit Boards. Resistors below 50 ohms act as a short circuit; above 100 ohms as an open circuit. The circuit is a simple multivibrator switched on by transistor T3. The components in the base of T3 are D1, R1, R2, and the test resistance. With a 1.5 volt supply, there is insufficient voltage to turn on a semiconductor connected to the test terminals.

SIMPLE CONTINUITY TESTER



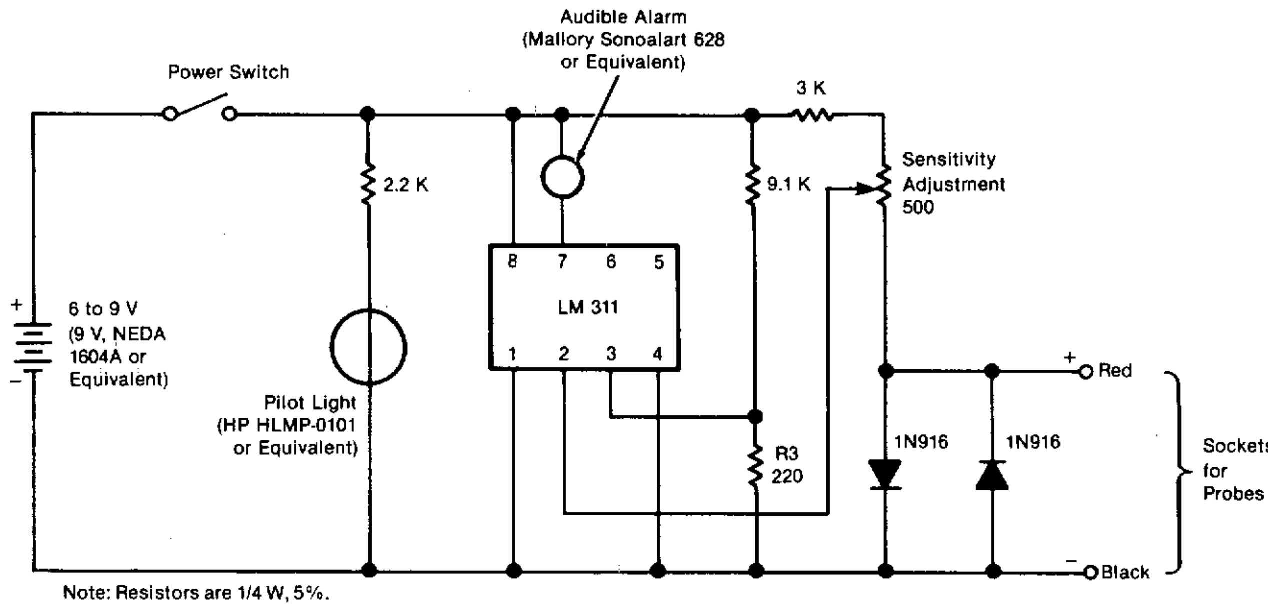
ELECTRONIC ENGINEERING

Fig. 84-3

Circuit Notes

The pitch of the tone is dependent upon the resistance under test. The tester will respond to resistance of hundreds of kilohms, yet it is possible to distinguish differences of just a few tens of ohms in low-resistance circuits. Q1 and Q2 form a multivibrator, the frequency of which is influenced by the resistance between the test points. The output stage Q3 and Q4 will drive a small loudspeaker or a telephone earpiece.

ADJUSTABLE, AUDIBLE CONTINUITY TESTER FOR DELICATE CIRCUITS



Note: Resistors are 1/4 W, 5%.

NASA

Fig. 84-4

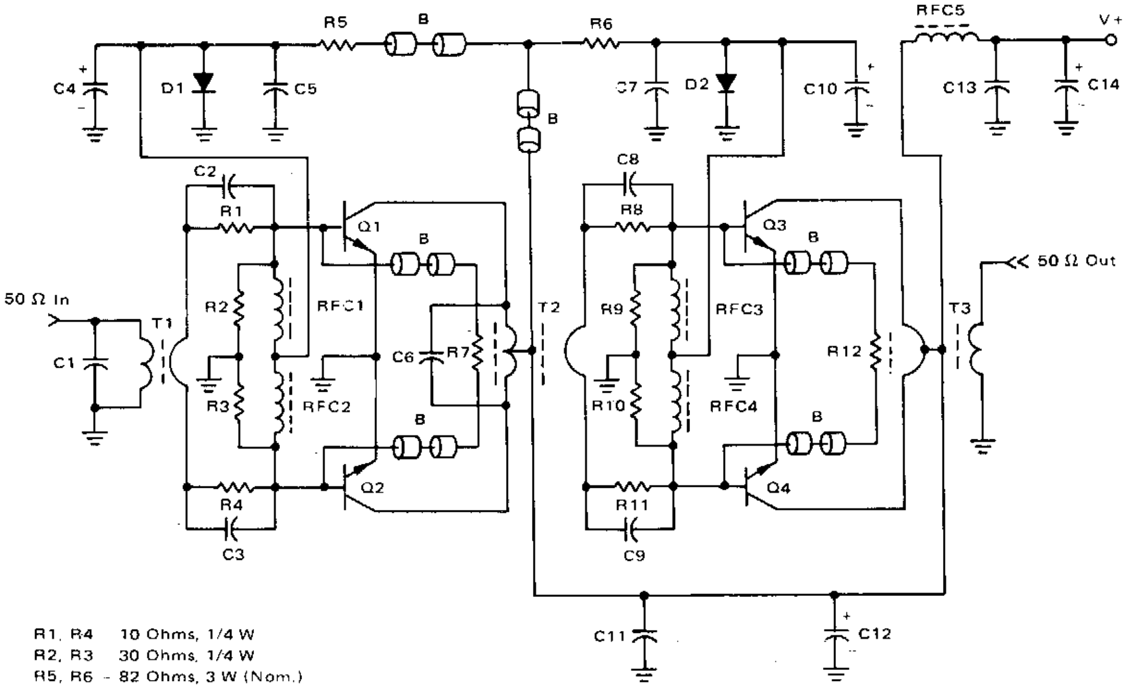
Circuit Notes

The tester gives an audible indication, making it unnecessary for the user to look directly at the instrument to observe a meter reading. In addition, the current and voltage of the tester are strictly limited. It can apply no more than 0.6 volts dc and no more than 3 mA through the probes. It can therefore be used safely on circuit boards in which semiconductor components have been installed, and on complementary metal oxide/semiconductor integrated circuits, which are highly susceptible to damage during testing. The tester can be adjusted to indicate continuity below any resistance value up to 35 ohms. For example, if the user sets the tester to 30 ohms, the unit will emit an audible tone whenever the resistance between the probes is 30 ohms or less; if, for example, the resistance is 30.2 ohms, the unit will remain silent.

85
RF Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW-DISTORTION 1.6 TO 30 MHz SSB DRIVER



R1, R4 10 Ohms, 1/4 W
R2, R3 30 Ohms, 1/4 W
R5, R6 – 82 Ohms, 3 W (Nom.)
R7 – 47 Ohms, 1/4 W
R8, R11 – 6.8 Ohms, 1/4 W
R9, R10 – 15 Ohms, 1/4 W
R12 – 130 Ohms, 1/4 W

C1 – 39 pF Dipped Mica
C2, C3 – 680 pF Ceramic Disc
C4, C10 – 220 μF, 4 V, Tantalum
C5, C7, C11, C13 – 0.1 μF Ceramic Disc
C6 – 56 pF Dipped Mica
C8, C9 – 1200 pF Ceramic Disc
C12, C14 – 10 μF, 25 V Tantalum

RFC5 – Ferroxcube VK200 19/4B
RFC1, 2, 3, 4 – 10 μH Molded Choke

B – Ferrite Beads (Fair-Rite Prod. Corp. #2643000101 or Ferroxcube #56 590 65/3B)

D1, D2 – 1N4001

Q1, Q2 – MRF476
Q3, Q4 – MRF475

T1, T2 – 4:1 Impedance Transformer
T3 – 1:4 Impedance Transformer

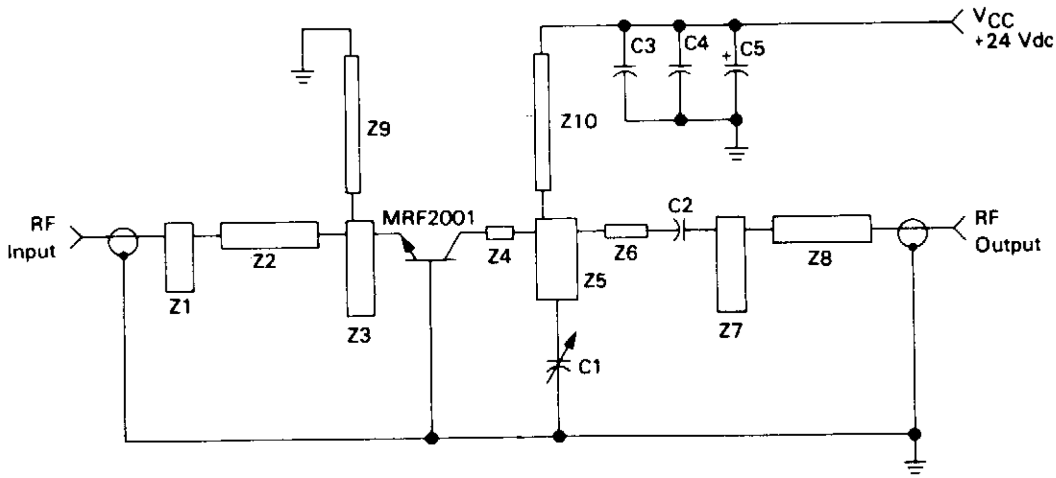
MOTOROLA

Fig. 85-1

Circuit Notes

The amplifier provides a total power gain of about 25 dB, and the construction technique allows the use of inexpensive components throughout. The MRF476 is specified as a 3 watt device and the MRF475 has an output power of 12 watts. Both are extremely tolerant to overdrive and load mismatches, even under CW conditions. Typical IMD numbers are better than −35 dB, and the power gains are 18 dB and 12 dB, respectively, at 30 MHz. The bias currents of each stage are individually adjustable with R5 and R6. Capacitors C4 and C10 function as audio-frequency bypasses to further reduce the source impedance at the frequencies of modulation. Gain leveling across the band is achieved with simple RC networks in series with the bases, in conjunction with negative feedback. The amplitude of the out-of-phase voltages at the bases is inversely proportional to the frequency as a result of the series inductance in the feedback loop and the increasing input impedance of the transistor at low frequencies. Conversely, the negative feedback lowers the effective input impedance presented to the source (not the input impedance of the device itself) and with proper voltage slope would equalize it. With this technique, it is possible to maintain an input VSWR of 1.5:1 or less than 1.6 to 30 MHz.

1 WATT, 2.3 GHz AMPLIFIER



C1 — 0.4-2.5 pF Johanson 7285*
C2, C3 — 68 pF, 50 mil ATC**
C4 — 0.1 μF, 50 V
C5 — 4.7 μF, 50 V Tantalum

Z1-Z10 — Microstrip; see Photomaster, Figure 3

Board Material — 0.0625″ 3M Glass Teflon,***
$\epsilon_r = 2.5 \pm 0.05$
*Johanson Manufacturing Corp., 400 Rockaway Valley Road, Boonton, NJ 07005
**American Technical Ceramics, One Norden Lane, Huntington Station, NY 11746
***Registered Trademark of Du Pont

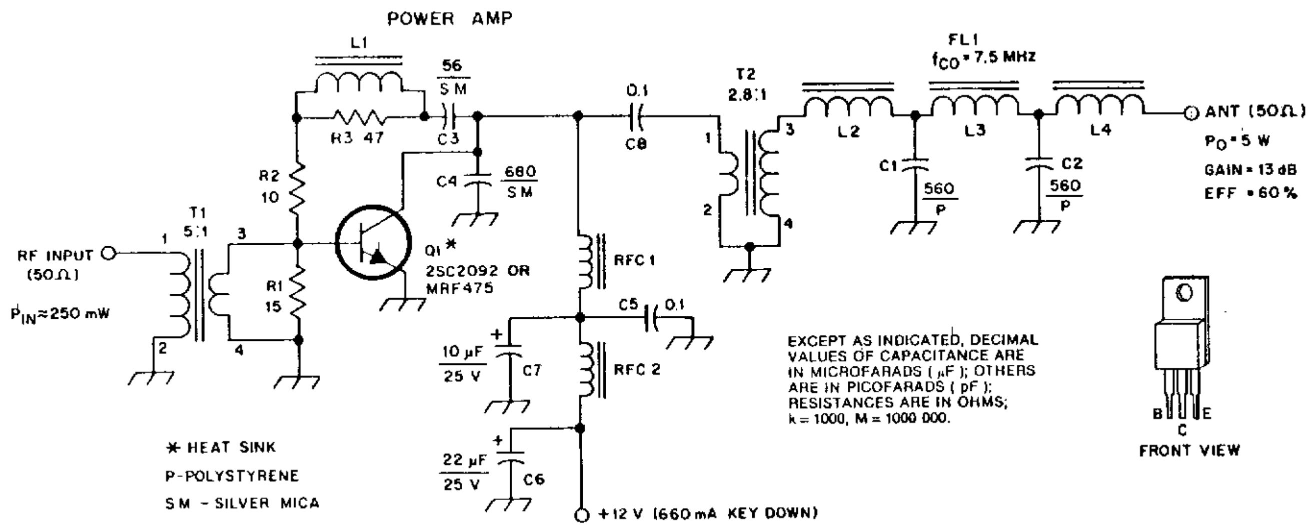
MOTOROLA

Fig. 85-2

Circuit Notes

Simplicity and repeatability are featured in this 1 watt S-band amplifier design. The design uses an MRF2001 transistor as a common base, Class C amplifier. The amplifier delivers 1 watt output with 8 dB minimum gain at 24 V, and is tunable from 2.25 to 2.35 GHz. Applications include microwave communications equipment and other systems requiring medium power, narrow band amplification. The amplifier circuitry consists almost entirely of distributed microstrip elements. A total of six additional components, including the MRF2001, are required to build a working amplifier. The input and output impedances of the transistor are matched to 50 ohms by double section low pass networks. The networks are designed to provide about 3% 1 dB power bandwidth while maintaining a collector efficiency of approximately 30%. There is one tuning adjustment in the amplifier—C1 in the output network. Ceramic chip capacitors, C2 and C3, are used for dc blocking and power supply decoupling. Additional low frequency decoupling is provided by capacitors C4 and C5.

5-W RF POWER AMPLIFIER



L1—0.22-µH inductor. Small RF choke or 8 turns of no. 24 enam wire on an Amidon T-37-6 toroid.
L2, L4—0.8-µH inductor. 12 turns of no. 24 enam wire on an Amidon T-50-2 toroid.
L3—1.67-µH inductor. 18 turns of no. 24 enam wire on an Amidon T-50-2 toroid.
RFC1—2.8 µH choke. 24 turns of no. 26 enam wire on an Amidon T-50-2 toroid.
RFC2—42 µH choke. 10 turns of no. 26 enam wire on an Amidon FT-37-43 toroid.
T1—Primary has 16 turns of no. 26 enam wire on an Amidon FT-37-43 toroid. Secondary has 6 turns of no. 26 enam wire.
T2—Primary (Q1 side) has 9 turns of no. 24 enam wire on an Amidon FT-50-43 toroid. Secondary has 15 turns of no. 24 enam wire.

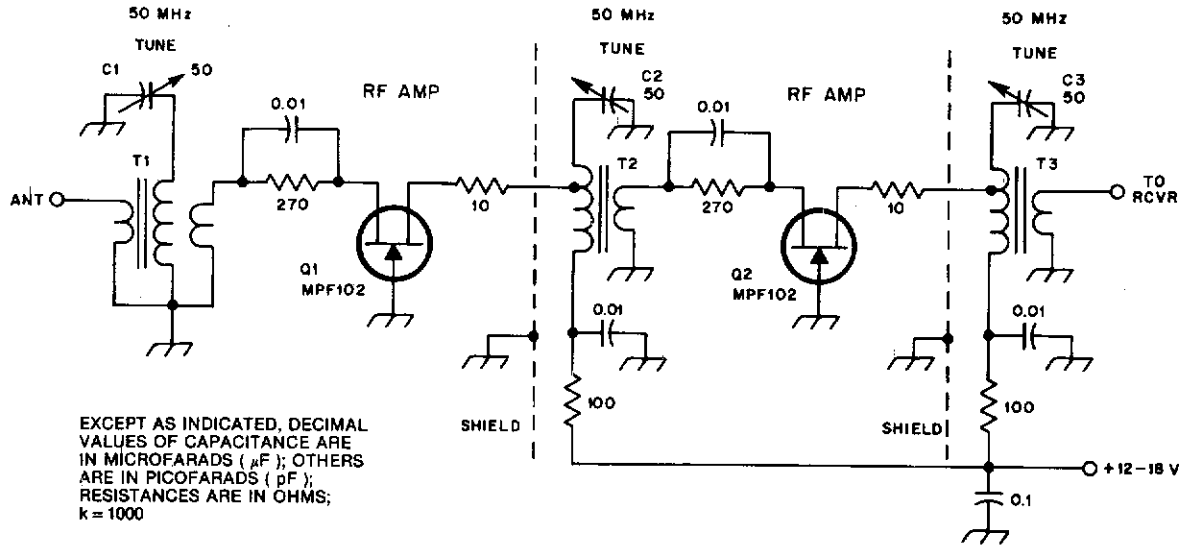
QST

Fig. 85-3

Circuit Notes

Numbered components are so designated for PC-board layout purposes. C5 and C8 are disc ceramic. C6 and C7 are tantalum or electrolytic. R1, R2 and R3 are ½ W carbon composition resistors. Silver-mica capacitors may be substituted for polystyrene (P) types. Impedance transformation ratios are shown above T1 and T2.

6-METER PREAMPLIFIER PROVIDES 20 dB GAIN AND LOW NF



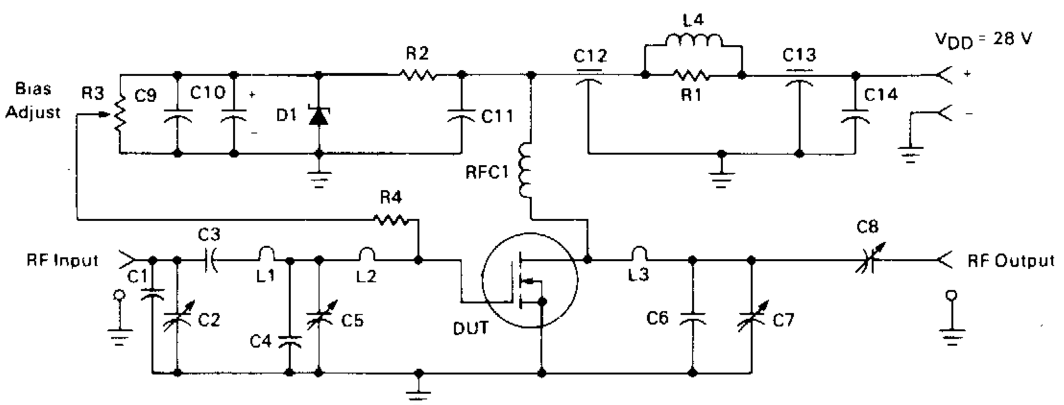
QST

Fig. 85-4

Circuit Notes

C1, C2, and C3 are miniature ceramic or plastic trimmers. T1 (main winding) is 0.34 μH. Use of 11 turns of no. 24 enameled wire on a T37-10 toroid core. The antenna winding has one turn, and Q1 the source winding has three turns. T2 primary consists of 11 turns of no. 24 enameled wire on a T37-10 toroid. Tap Q1 drain is three turns from C2 the end of the winding. The secondary has three turns. T3 is the same as T2, except its secondary has one turn.

125 WATT 150 MHz AMPLIFIER



C1 — 35 pF Unleco
C2, C5 — Arco 462, 5-80 pF
C3 — 100 pF Unleco
C4 — 25 pF Unleco
C6 — 40 pF Unleco
C7 — Arco 461, 2.7-30 pF
C8 — Arco 463, 9-180 pF
C9, C11, C14 — 0.1 μF Erie Redcap
C10 — 50 μF, 50 V
C12, C13 — 680 pF Feedthru
D1 — 1N5925A Motorola Zener

L1 — #16 AWG, 1-1/4 Turns, 0.213" ID
L2 — #16 AWG, Hairpin 0.25" 0.062"
L3 — #14 AWG, Hairpin 0.47" 0.2"
L4 — 10 Turns #16 AWG Enameled Wire on R1
RFC1 — 18 Turns #16 AWG Enameled Wire, 0.3" ID
R1 — 10 Ω, 2.0 W
R2 — 1.8 kΩ, 1/2 W
R3 — 10 kΩ, 10 Turn Bourns
R4 — 10 kΩ, 1/4 W

MOTOROLA *Fig. 85-5*

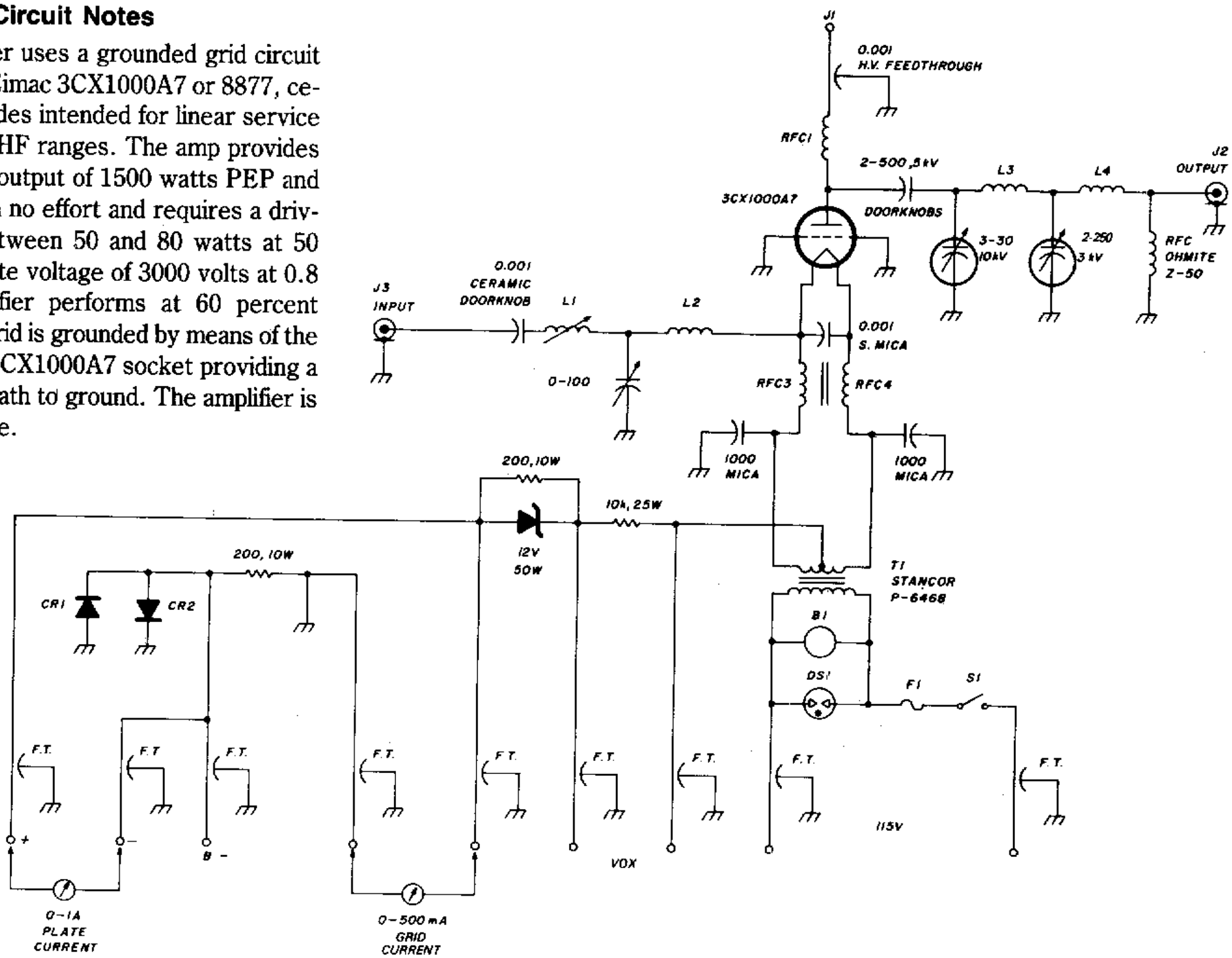
Circuit Notes

This amplifier operates from a 28 Vdc supply. It has a typical gain of 12 dB, and can survive operation into a 30:1 VSWR load at any phase angle with no damage. The amplifier has an AGC range in excess of 20 dB. This means that with input power held constant at the level that provides 125 watts output, the output power may be reduced to less than 1.0 watt continuously by driving the dc gate voltage negative from its I_{DQ} value.

6-METER KILOWATT AMPLIFIER

Circuit Notes

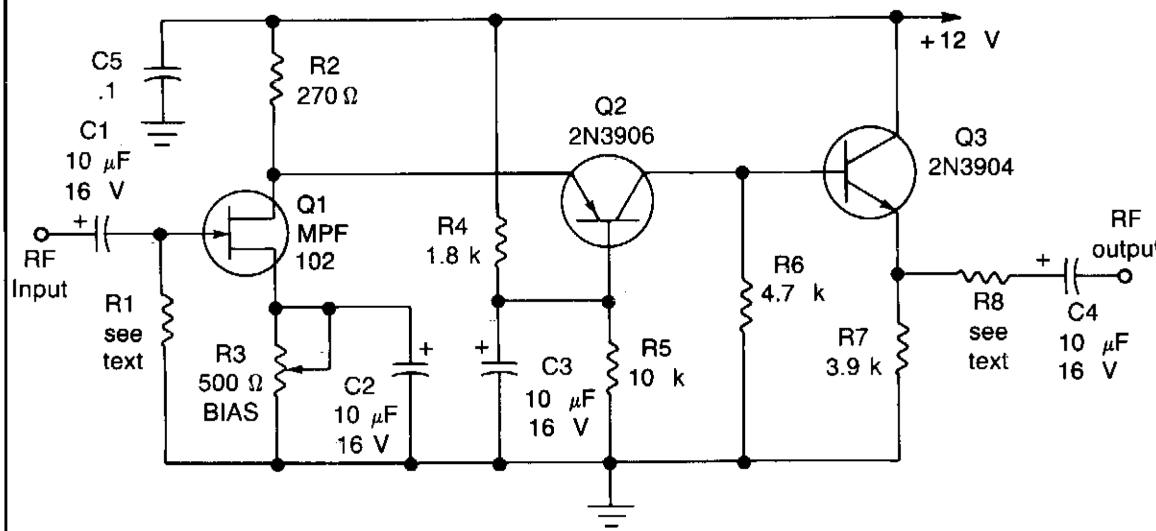
The amplifier uses a grounded grid circuit with either the Eimac 3CX1000A7 or 8877, ceramic/metal triodes intended for linear service in the HF and VHF ranges. The amp provides the legal power output of 1500 watts PEP and CW service with no effort and requires a driver delivering between 50 and 80 watts at 50 MHz. With a plate voltage of 3000 volts at 0.8 amps the amplifier performs at 60 percent efficiency. The grid is grounded by means of the grid ring of the 3CX1000A7 socket providing a low-inductance path to ground. The amplifier is completely stable.



HAM RADIO

Fig. 85-6

BROADCAST-BAND RF AMPLIFIER



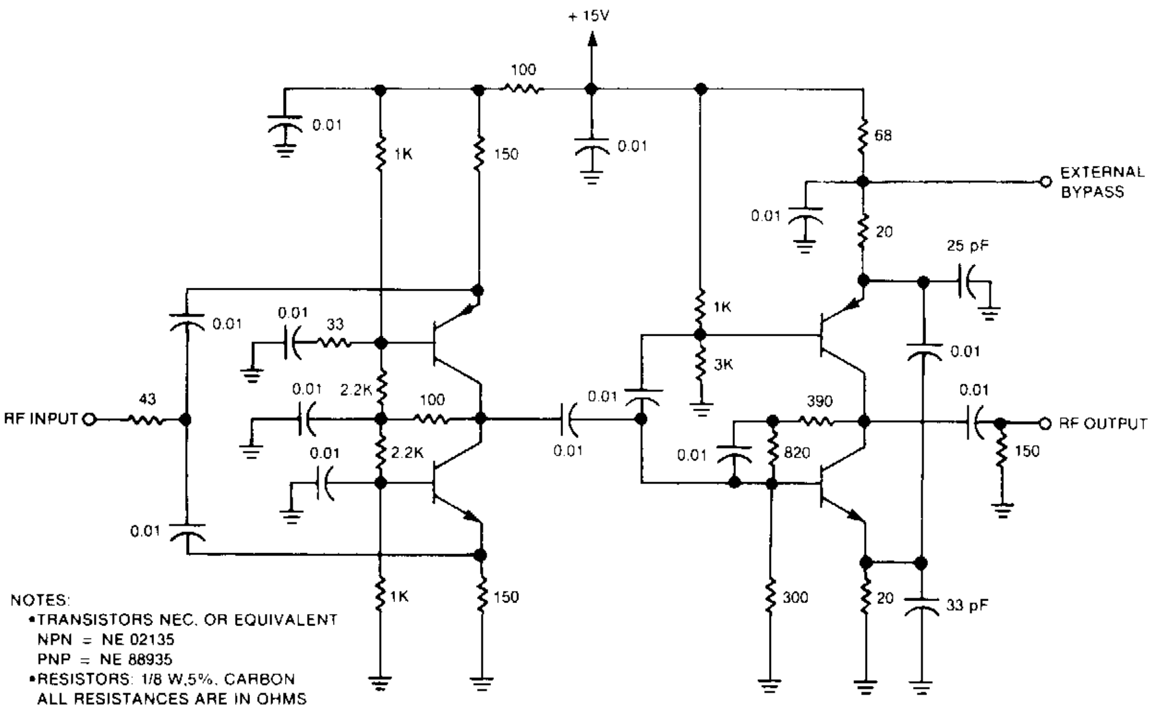
RADIO-ELECTRONICS

Fig. 85-7

Circuit Notes

The circuit has a frequency response ranging from 100 Hz to 3 MHz; gain is about 30 dB. Field-effect transistor Q1 is configured in the common-source self-biased mode; optional resistor R1 sets the input impedance to any desired value. Commonly, it will be 50 ohms. The signal is then direct-coupled to Q2, a common-base circuit that isolates the input and output stages and provides the amplifier's exceptional stability. Q3 functions as an emitter follower, to provide low output impedance (about 50 ohms). For higher output impedance, include resistor R8. It will affect impedance according to this formula: $R8 \sim R_{out} - 50$. Otherwise, connect output capacitor C4 directly to the emitter of Q3.

IMPROVED RF ISOLATION AMPLIFIER



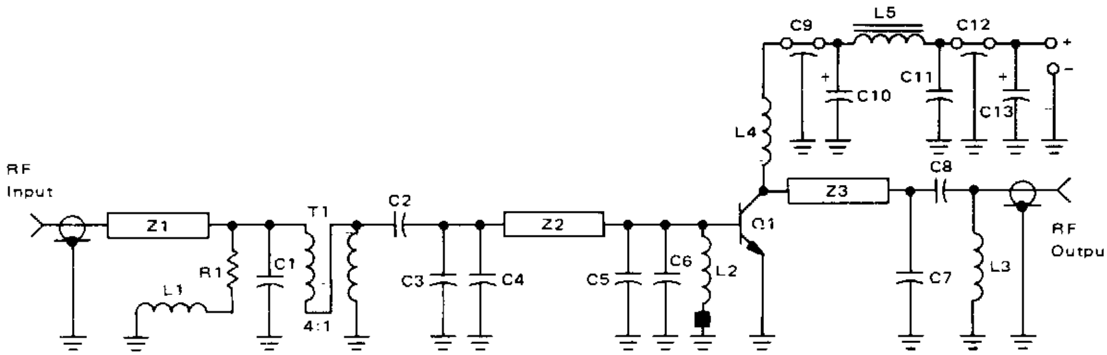
NOTES:
- TRANSISTORS NEC. OR EQUIVALENT
  NPN = NE 02135
  PNP = NE 88935
- RESISTORS: 1/8 W,5%, CARBON
  ALL RESISTANCES ARE IN OHMS
  UNLESS OTHERWISE NOTED.
- ALL CAPACITANCES ARE IN MICROFARADS
  UNLESS OTHERWISE NOTED.

NASA *Fig. 85-8*

Circuit Notes

This wideband RF isolation amplifier has a frequency response of 0.5 to 400 MHz ± 0.5 dB. This two stage amplifier can be used in applications requiring high reverse isolation, such as receiver intermediate-frequency (IF) strips and frequency distribution systems. Both stages use complementary-symmetry transistor arrangements. The input stage is a common-base connection for the complementary circuit. The output stage, which supplies the positive gain, is a common-emitter circuit using emitter degeneration and collector-base feedback for impedance control.

A 10 WATT 225-400 MHz AMPLIFIER



C1 – 8.2 pF Chip*
C2 – 270 pF Chip*
C3 – 36 pF Chip*
C4, C7 – 15 pF Chip*
C5, C6 – 50 pF Chip*
C8 – 82 pF Chip*
C9, C12 – 680 pF Feedthru
C10, C13 – 1.0 µF 50 V Tantalum
C11 – 0.1 µF Erie Redcap
L1, L3 – 3 Turns #22 AWG 1/8″ (3.175 mm) ID
L2 – 0.15 µH Molded Choke
L4 – 0.15 µH Molded Choke with Ferroxcube Bead (Ferroxcube 56 590-65/4B on Ground End of Choke)
L5 – Ferroxcube VK200-19/4B

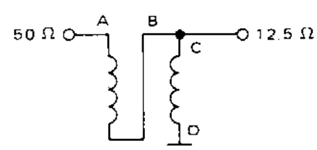
*100 mil A.C.I. Chip Capacitors

R1 – 36 Ω 1/4 Watt
T1 – 25 Ω Subminiature Coax (Type UT34 25) – 1.75 inches (44.45 mm) long
Z1 – Microstrip Line
720 mils L X 162 mils W
18.29 mm L X 4.115 mm W
Z2 – Microstrip Line
680 mils L X 162 mils W
17.27 mm L X 4.115 mm W
Z3 – Microstrip Line
2200 mils L X 50 mils W
55.88 mm L X 1.27 mm W
Board – 0.0625″ (1.588 mm) Glass Teflon, ϵ_r = 2.56
Q1 – MRF331

MOTOROLA

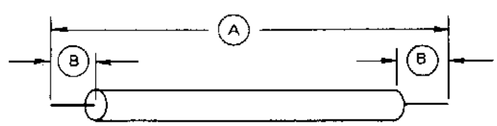
Fig. 85-9

SCHEMATIC REPRESENTATION



ASSEMBLY AND PICTORIAL

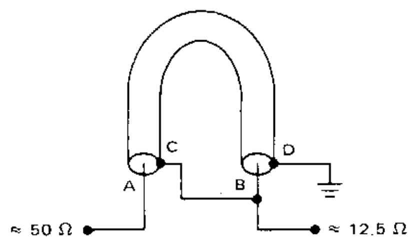
Transformer Dimensions (not to scale)



(A) – 1.75 inches (4.445 cm)
(B) – 0.1875 inch (0.476 cm)

A 10 WATT 225-400 MHz AMPLIFIER, Continued.

Transformer Connections



Circuit Notes

This broadband amplifier covers the 225-400 MHz military communications band producing 10 watt RF output power and operating from a 28 volt supply. The amplifier can be used as a driver for higher power devices such as 2N6439 and MRF327. The circuit is designed to be driven by a 50 ohm source and operate into a nominal 50 ohm load. The input matching network consists of a section composed of C3, C4, Z2, C5 and C6. C2 is a dc blocking capacitor, and T1 is a 4:1 impedance ratio coaxial transformer. Z1 is a 50 ohm transmission line. A compensation network consisting of R1, C1, and L1 is used to improve the input VSWR and flatten the gain response of the amplifier. L2 and a small ferrite bead make up the base bias choke. The output network is made up of a microstrip L-section consisting of Z3 and C7, and a high pass section consisting of C8 and L3. C8 also serves as a dc blocking capacitor. Collector decoupling is accomplished through the use of L4, L5, C9, C10, C11, C12, and C13.

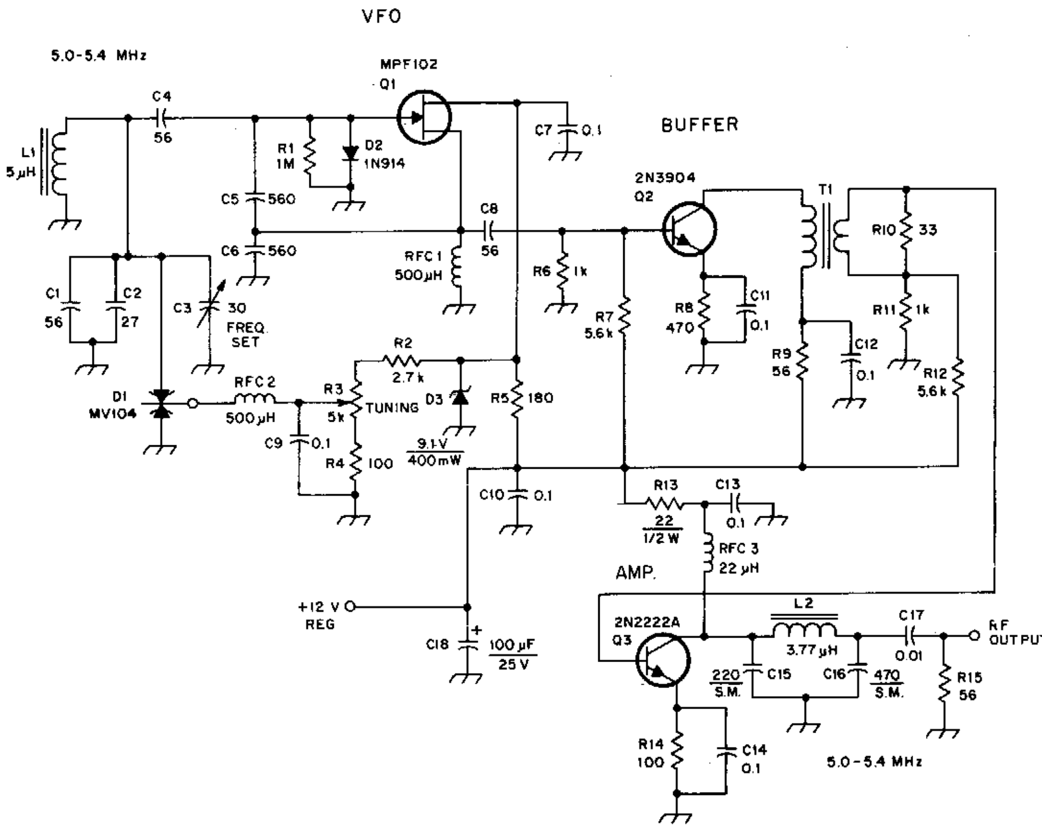
86

RF Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

5 MHz VFO

5 MHz VFO



QST

Fig. 86-1

Circuit Notes

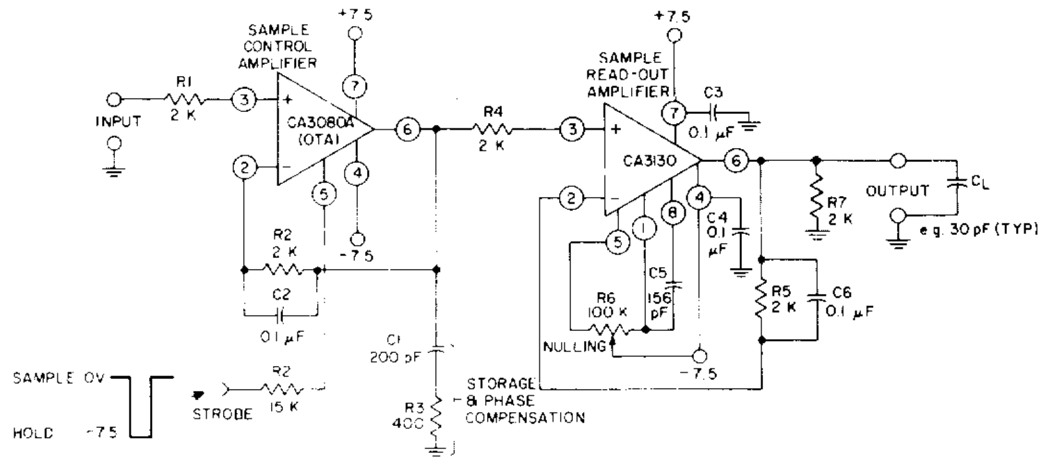
A JFET (Q1) serves as the oscillator. D2 helps to stabilize the transistor by limiting positive sinewave peaks and stabilizing the bias. Output from Q1 is supplied to a class A buffer, Q2. It operates as a broadband amplifier by means of T1, which is untuned. Output amplifier Q3 is also a class A stage. A low-pass, single-section filter is used at the output of Q3 to remove some of the harmonic currents generated within the system. The filter output impedance is 50 ohms. The injection level to the mixer is 600 mV p-p.

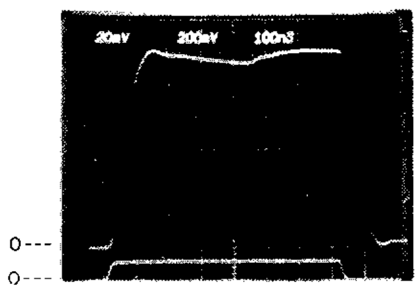
87

Sample-and-Hold Circuits

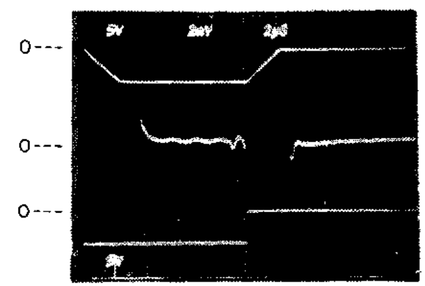
The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SAMPLE-AND-HOLD CIRCUIT





TOP TRACE: OUTPUT — 20 mV/DIV. & 100 ns/DIV.
BOTTOM TRACE: INPUT—200 mV/DIV. & 100 ns/DIV.



TOP TRACE: OUTPUT—5 V/DIV. & 2 μs/DIV.
CENTER TRACE: DIFFERENTIAL COMPARISON OF INPUT & OUTPUT—2 mV/DIV. & 2 μs/DIV.
BOTTOM TRACE: INPUT—5 V/DIV. & 2 μs/DIV.

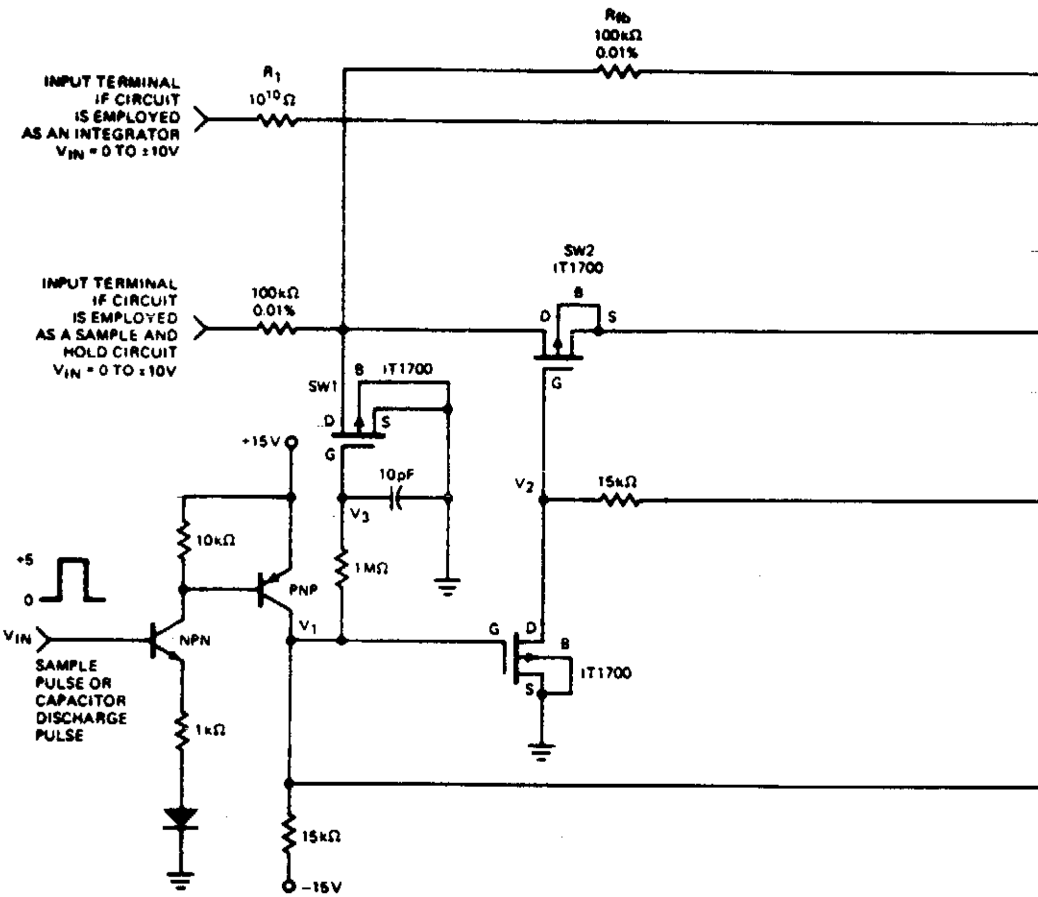
GENERAL ELECTRIC /RCA

Fig. 87-1

Circuit Notes

The circuit uses a CA3130 BiMOS op amp as the sample-readout amplifier for the storage (sample-holding) capacitor C1, and a CA3080A as the sample-control amplifier. Applications in linear systems to temporarily store analog data include DVM systems, industrial process-control, multiplex systems, and A-D converters.

SAMPLE AND HOLD CIRCUIT II

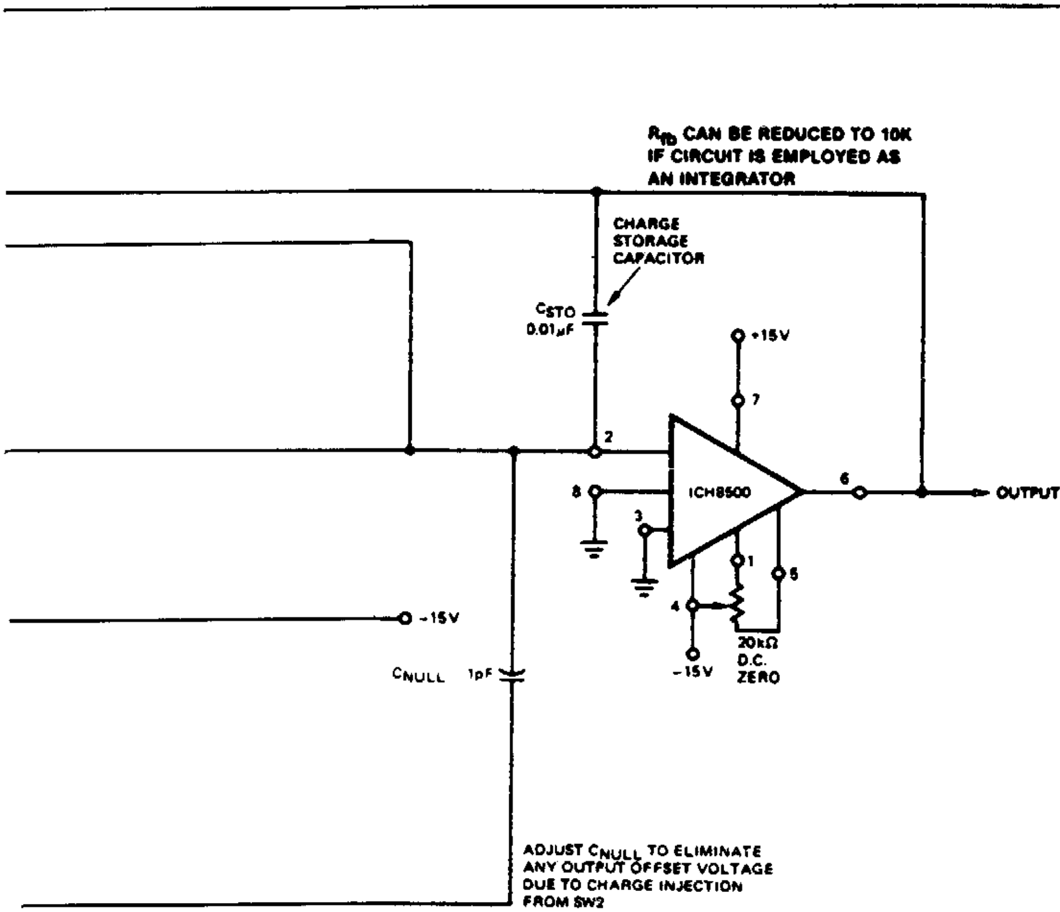


INTESIL

Fig. 87-2

Circuit Notes

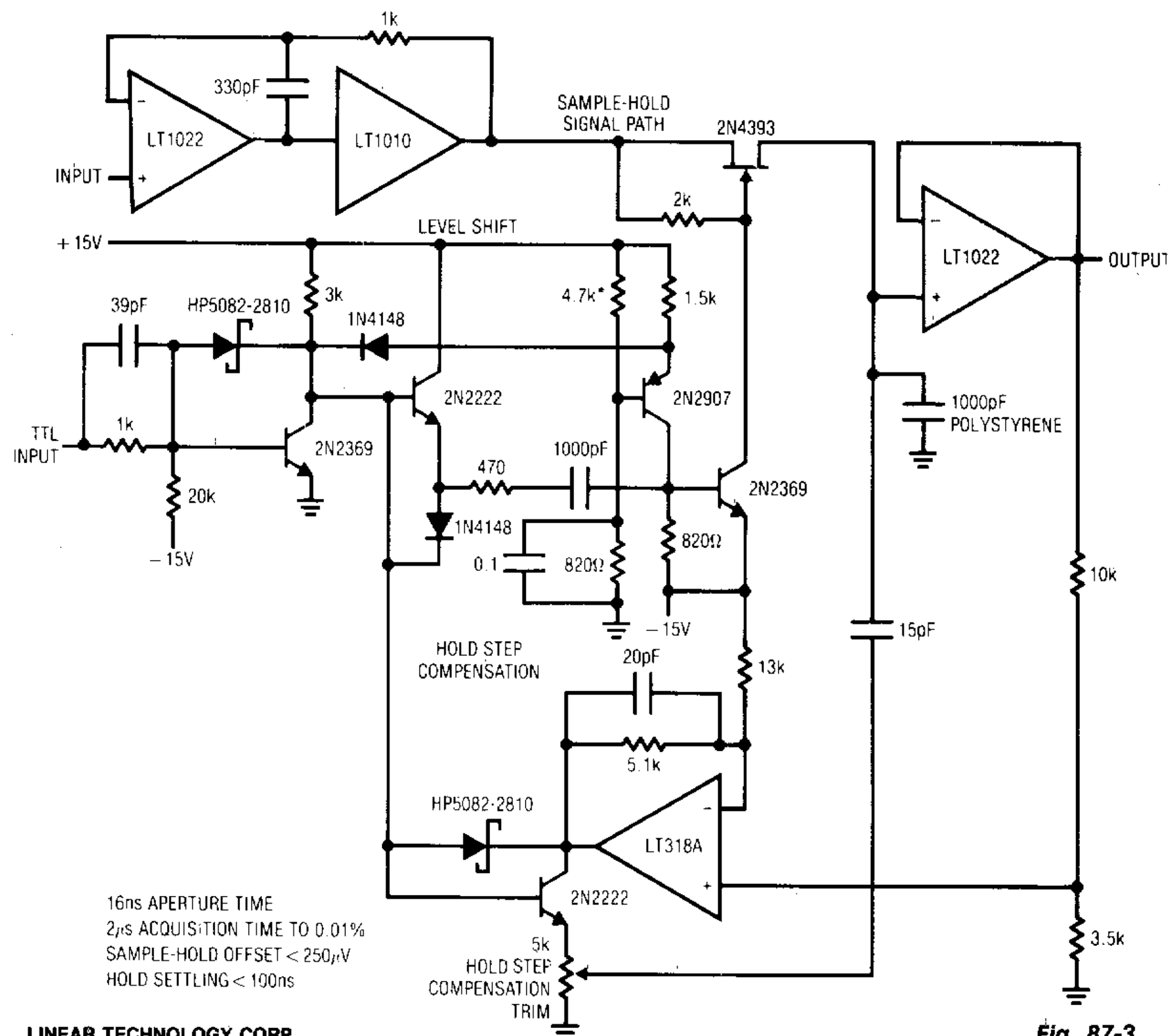
This circuit rapidly charges capacitor C_{STO} to a voltage equal to an input signal. The input signal is then electrically disconnected from the capacitor with the charge still remaining on C_{STO}. Since C_{STO} is in the negative feedback loop of the operational amplifier, the output voltage of the amplifier is equal to the voltage across the capacitor. Ideally, the voltage across C_{STO} should remain constant causing the output of the amplifier to remain constant as well. However, the voltage across C_{STO} will decay at a rate proportional to the current being injected or taken out of the current summing node of the amplifier. This current can come from four sources: leakage resistance of



C_{STO}, leakage current due to the solid state switch SW2, currents due to high resistance paths on the circuit fixture, and most important, bias current of the operational amplifier. If the ICH8500A operational amplifier is employed, this bias current is almost non-existent (less than 0.01pA). Note that the voltages on the source, drain and gate of switch SW2 are zero or near zero when the circuit is in the hold mode. Careful construction will eliminate stray resistance paths and capacitor resistance can be eliminated if a quality capacitor is selected. The net result is a low drift sample and hold circuit.

The circuit can double as an integrator. In this application the input voltage is applied to the integrator input terminal. The time constant of the circuit is the product of R1 and C_{STO}.

FAST, PRECISION SAMPLE-HOLD

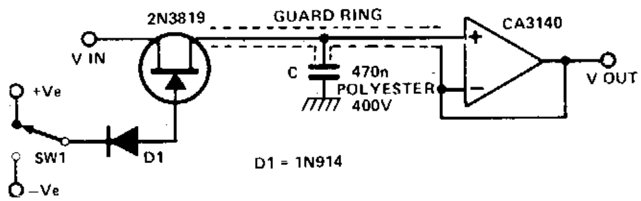


16ns APERTURE TIME
2μs ACQUISITION TIME TO 0.01%
SAMPLE-HOLD OFFSET < 250μV
HOLD SETTLING < 100ns

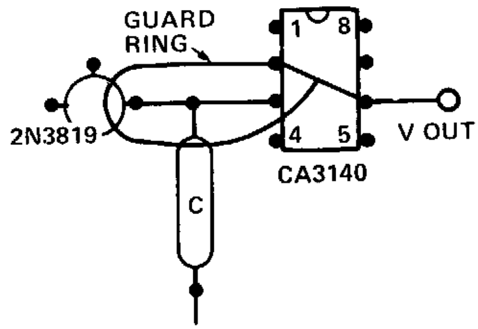
LINEAR TECHNOLOGY CORP.

Fig. 87-3

HIGH PERFORMANCE SAMPLE AND HOLD



PRINTED CIRCUIT BOARD LAYOUT



ELECTRONICS TODAY INTERNATIONAL

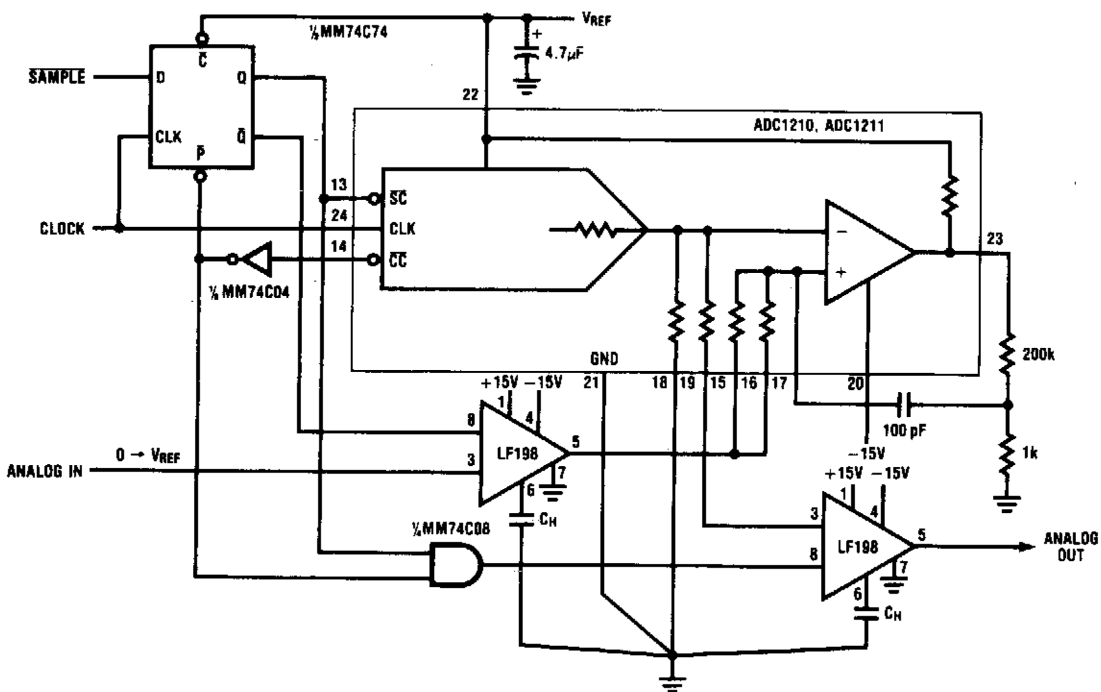
Fig. 87-4

Circuit Notes

When switch SW1 is positive, the FET is turned on, and has a resistance of about 400 ohm. The input voltage charges up the capacitor through the FET. When SW1 is negative, the FET is turned off (pinched off). To get a long storage time, the op amp must have a very low input bias current. For the CA3140, this current is about 10 pico amps. The rate at which the capacitor will be discharged by this current is based on the equation, C (dv/dt) = i where dv/dt is the rate of change of voltage on the capacitor. Therefore:

$$\frac{dv}{dt} = \frac{i}{C} = \frac{10^{-11}}{0.47 \times 10^{-6}} = 22\mu V/s$$

INFINITE SAMPLE AND HOLD AMPLIFIER



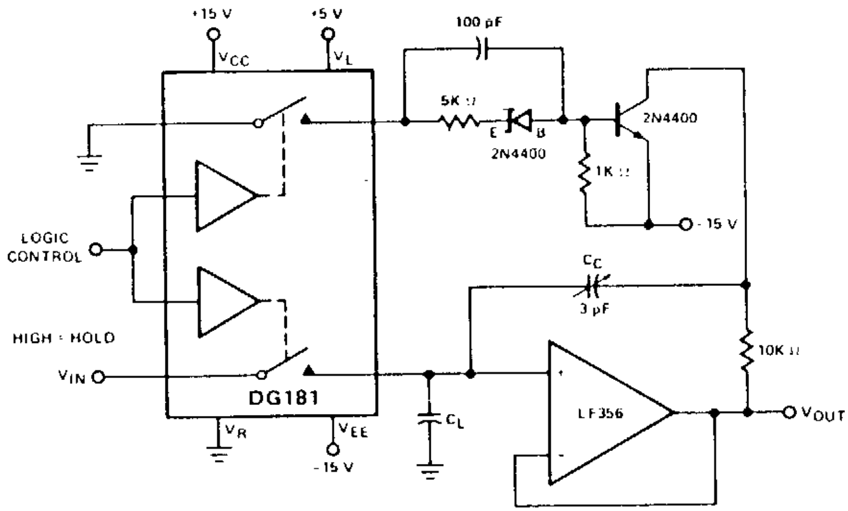
NATIONAL SEMICONDUCTOR CORP.

Fig. 87-5

Circuit Notes

During normal "hold" mode, the replicated analog voltage is buffered straight through the S/H amplifier to the output. Upon issuance of a $\overline{\text{SAMPLE}}$ signal, the S/H amplifier is placed in the hold mode, holding the voltage until the new analog voltage is valid. The same $\overline{\text{SAMPLE}}$ signal triggers an update to the input sample-and-hold amplifier. The most current analog voltage is captured and held for conversion. The previously determined voltage is held stable at the output during the conversion cycle while the SAR/D-to-A converter continuously adjusts to replicate the new input voltage. At the end of the conversion, the output sample-and-hold amplifier is once again placed in the track mode. The new analog voltage is then regenerated.

CHARGE COMPENSATED SAMPLE AND HOLD



(< 5 mV of Sample to Hold Offset when C_L = 1000 pF)

SILICONIX *Fig. 87-6*

Circuit Notes

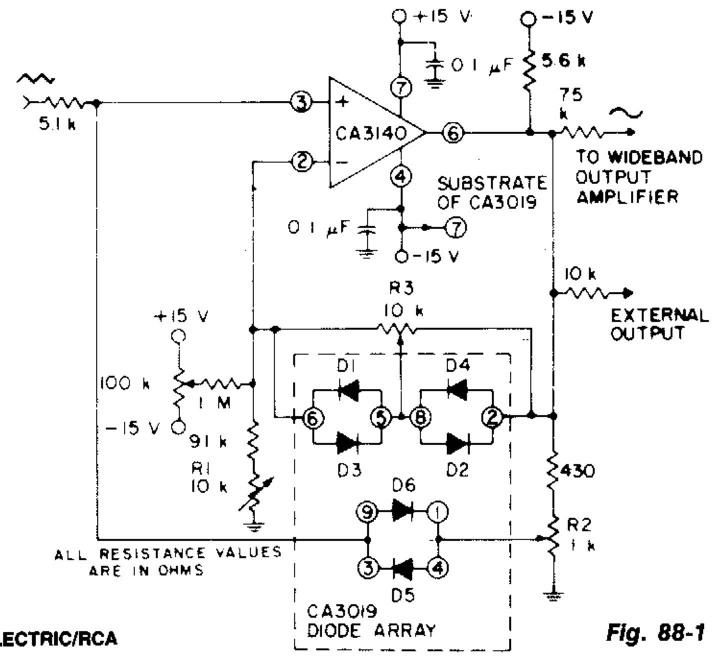
Less than ±5 pC charge transfer (less than 5 mV sample-to-hold offset when C_L = 1000 pF).

88

Sine-Wave Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SINE-WAVE SHAPER



GENERAL ELECTRIC/RCA

Fig. 88-1

Circuit Notes

Uses a CA3140 BiMOS op amp as voltage follower, together with diodes from a CA3019 array, to convert a triangular signal (such as obtained from a function generator) to a sine-wave output with typical THD less than 2%.

LOW DISTORTION SINE-WAVE OSCILLATOR

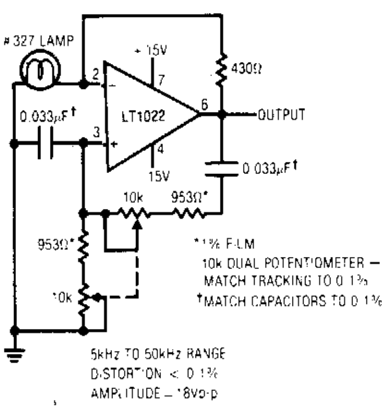
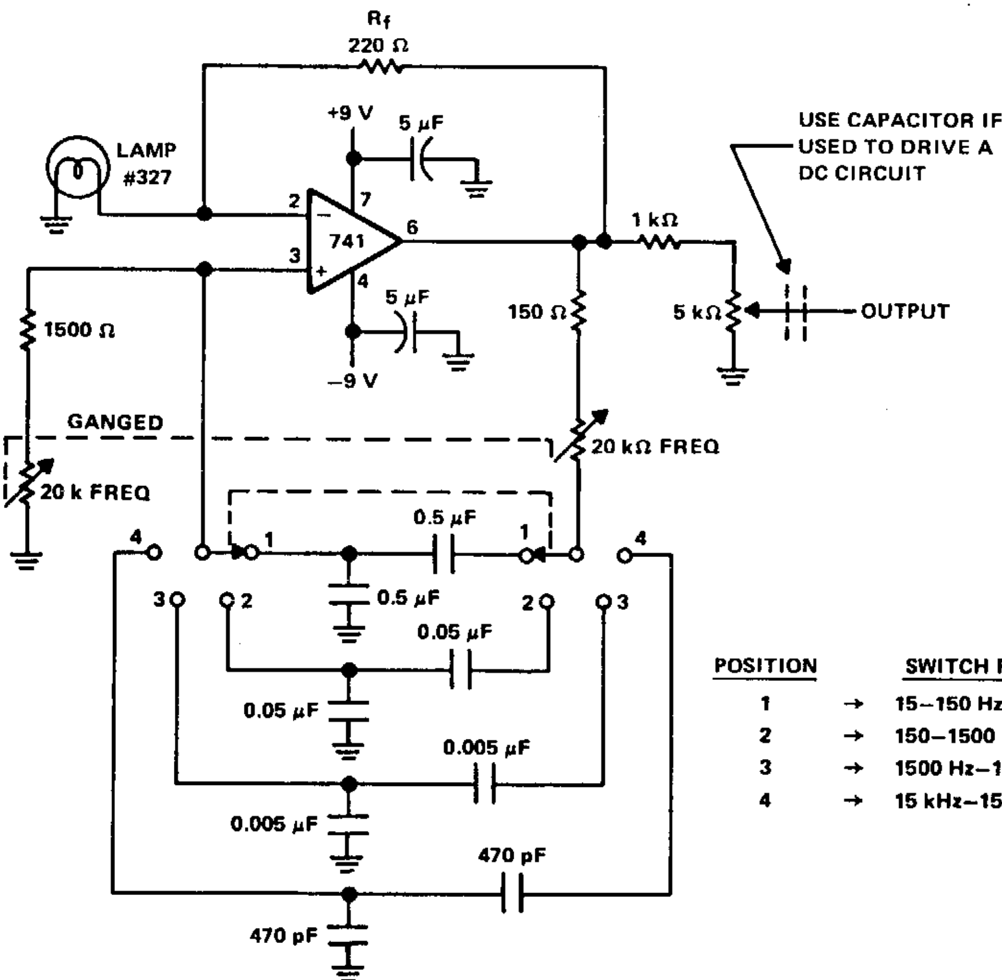


Fig. 88-2

LINEAR TECHNOLOGY CORP.

AUDIO OSCILLATOR



POSITION		SWITCH FREQ
1	→	15–150 Hz
2	→	150–1500 Hz
3	→	1500 Hz–15 kHz
4	→	15 kHz–150 kHz

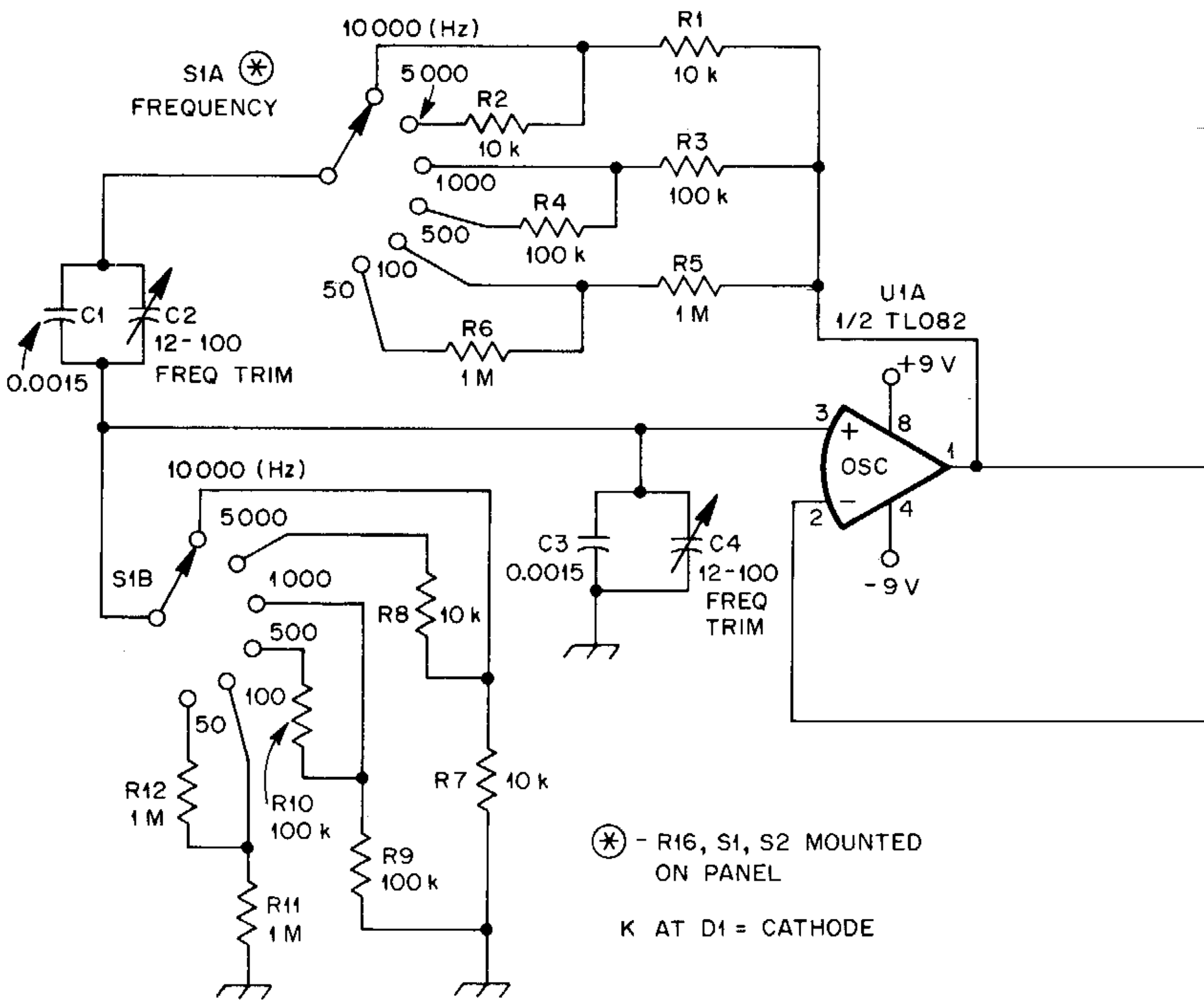
TEXAS INSTRUMENTS

Fig. 88-3

Circuit Notes

A Wien bridge oscillator produces sine waves with very low distortion level. The Wien bridge oscillator produces zero phase shift at only one frequency (f = ½ π RC) which will be the oscillation frequency. Stable oscillation can occur only if the loop gain remains at unity at the oscillation frequency. The circuit achieves this control by using the positive temperature coefficient of a small lamp to regulate gain (R_f/R_{LAMP}) as the oscillator attempts to vary its output. The oscillator shown here has four frequency bands covering about 15 Hz to 150 kHz. The frequency is continuously variable within each frequency range with ganged 20 k ohm potentiometers. The oscillator draws only about 4.0 mA from the 9-V batteries. Its output is from 4 to 5 V with a 10 k ohm load and the R_f (feedback resistor) is set at about 5% below the point of clipping. As shown, the center arm of the 5 k ohm output potentiometer is the output terminal. To couple the oscillator to a dc type circuit, a capacitor should be inserted in series with the output lead.

SIMPLE AUDIO SINE-WAVE GENERATOR

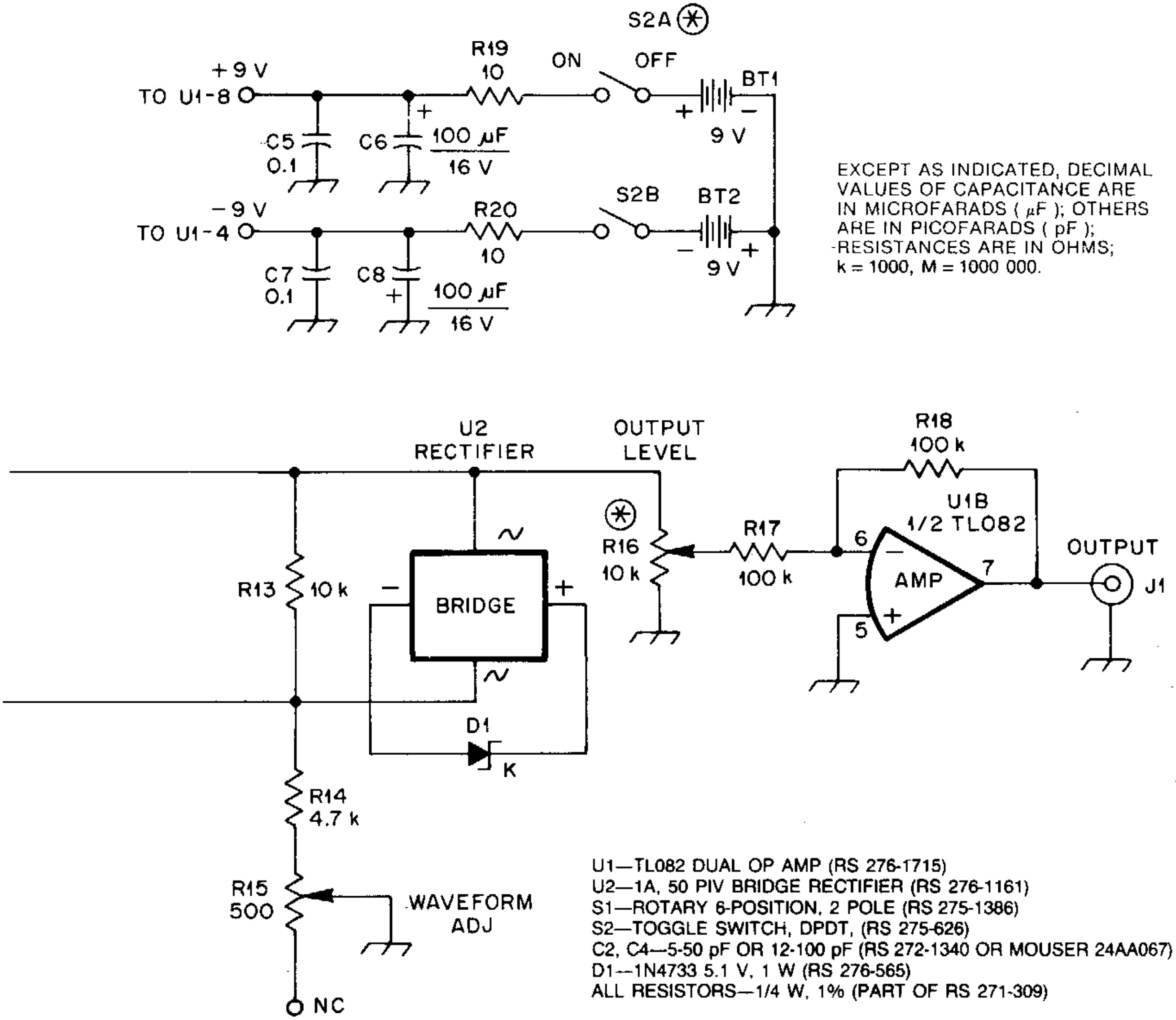


QST

Fig. 88-4

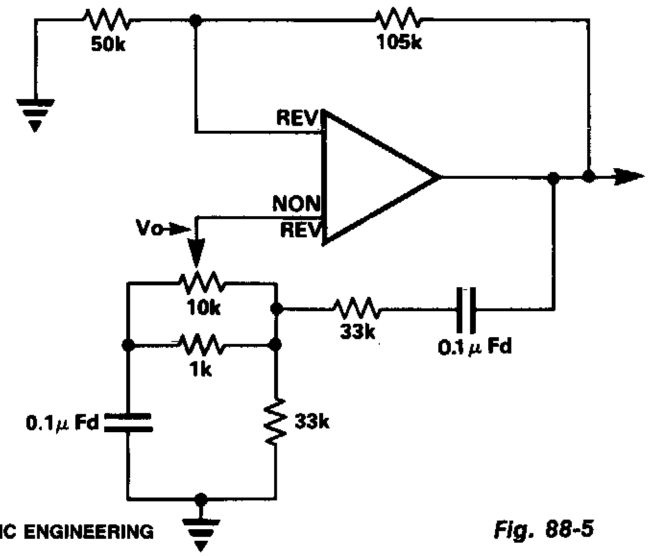
Circuit Notes

U1A, an op amp, oscillates at the frequency at which the phase shift in the Wien bridge network is exactly zero degrees. Changing bridge component values changes the oscillator frequency. In this circuit, we need change only the two resistors to do this. S1A chooses a value among R1 through R6, and S1B similarly selects a value from R7 through R12. U1A must provide enough gain to overcome losses in the bridge, but not so much gain that oscillation builds up to the point of overload and distortion. U2 and



C1 automatically regulate circuit gain to maintain oscillation. U2 places D1 across R13 with the proper polarity on both positive and negative alterations of the signal at pin 1 of U1. As the voltage at pin 1 of U1 approaches its peak value, D1 enters its Zener breakdown region, effectively shunting R13 with a resistive load. This increases the amount of negative feedback around U1, reducing its gain. R15, WAVEFORM ADJ, allows you to optimize circuit operation for lowest distortion. U1B provides isolation between oscillator and load. With the values shown for R17 and R18, U1B operates at unity gain.

LOW COST WIEN BRIDGE OSCILLATOR



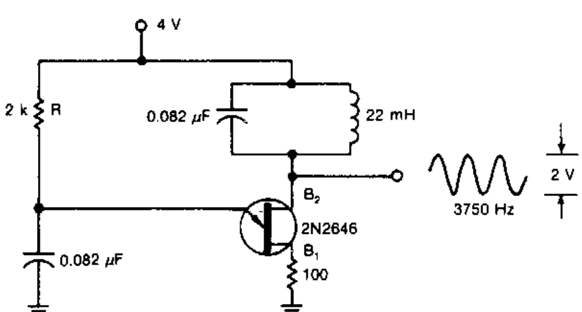
ELECTRONIC ENGINEERING

Fig. 88-5

Circuit Notes

In the circuit the frequency trimming component is arranged so that the voltage across it is in quadrature with the voltage V_0 from the bridge so that as it is adjusted the attenuation of the bridge only changes a little, avoiding the need for a two gang component. The range of variation of frequency is very limited. By using a high gain amplifier and metal film feedback resistors the loop gain can be set so that the unit just oscillates and the use of an automatic gain setting component, a thermistor for example, is eliminated.

MODIFIED UJT RELAXATION OSCILLATOR PRODUCES CLEAN AUDIO SINUSOIDS



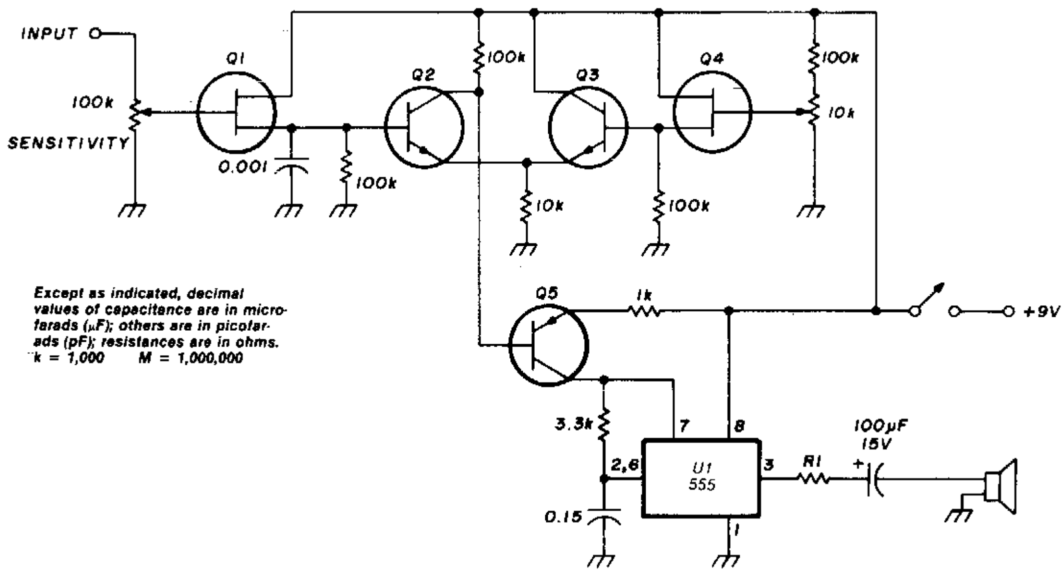
Circuit Notes

By placing a tuned circuit in the UJT oscillator's current-pulse path, a 3750-Hz sinusoid can be created at B2 with the component values shown.

ELECTRONIC DESIGN

Fig. 88-6

A 555 USED AS AN RC AUDIO OSCILLATOR



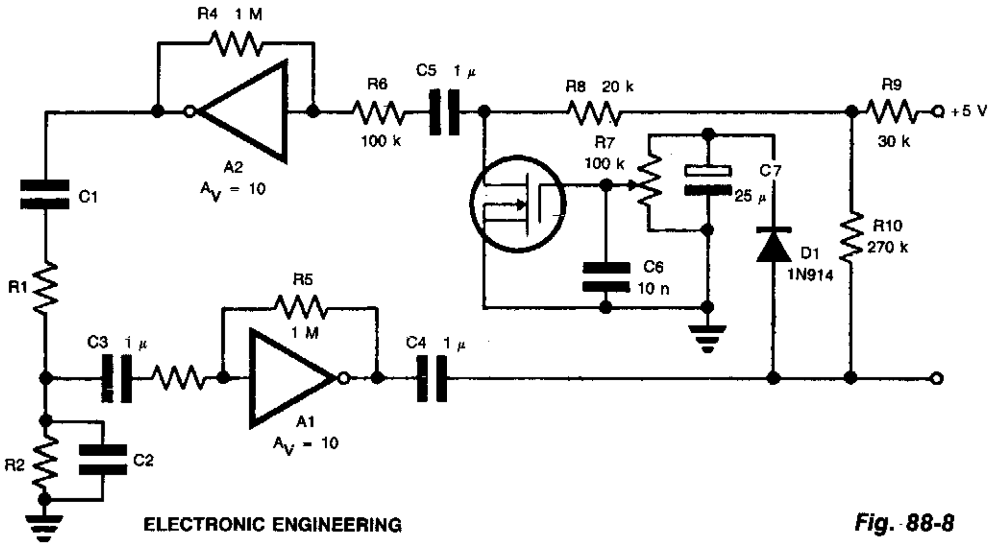
HAM RADIO

Fig. 88-7

Circuit Notes

Transistor Q5 and the 1000 ohm resistor form the variable element needed for controlling the frequency of VCO by limiting the charging current flowing into the 0.15 μF timing capacitor according to the forward bias being applied to Q5. As the voltage on pins 2 and 6 of U1 reach ⅔ V_{CC} (about 6 volts with a 9-volt supply) the timer will fire and pin 3 will be pulled low. Pin 7, an open collector output, goes low and begins to discharge the timing capacitor—through the 3.3 kilohm resistor. The discharge time provided by this resistor assures a reasonable, although asymmetrical, waveform for the aural signal generated by U1. At ⅓ V_{CC} the internal flip-flop resets, the output on pin 3 goes high, the open collector output on pin 7 floats, and the timing cycle begins again.

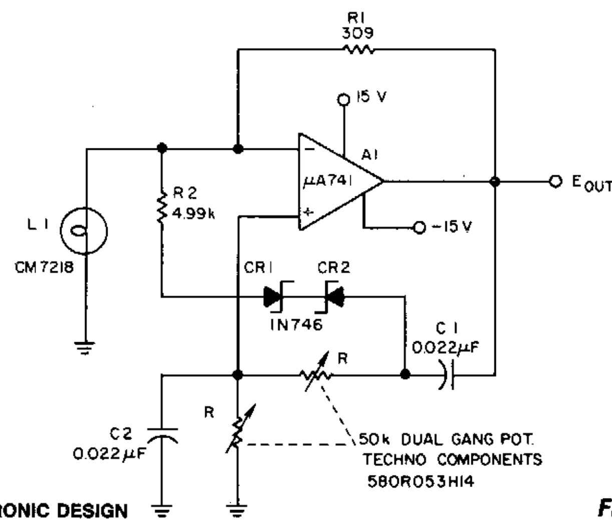
WIEN BRIDGE OSCILLATOR USES CMOS CHIP



ELECTRONIC ENGINEERING

Fig. 88-8

ADJUSTABLE SINE-WAVE AUDIO OSCILLATOR



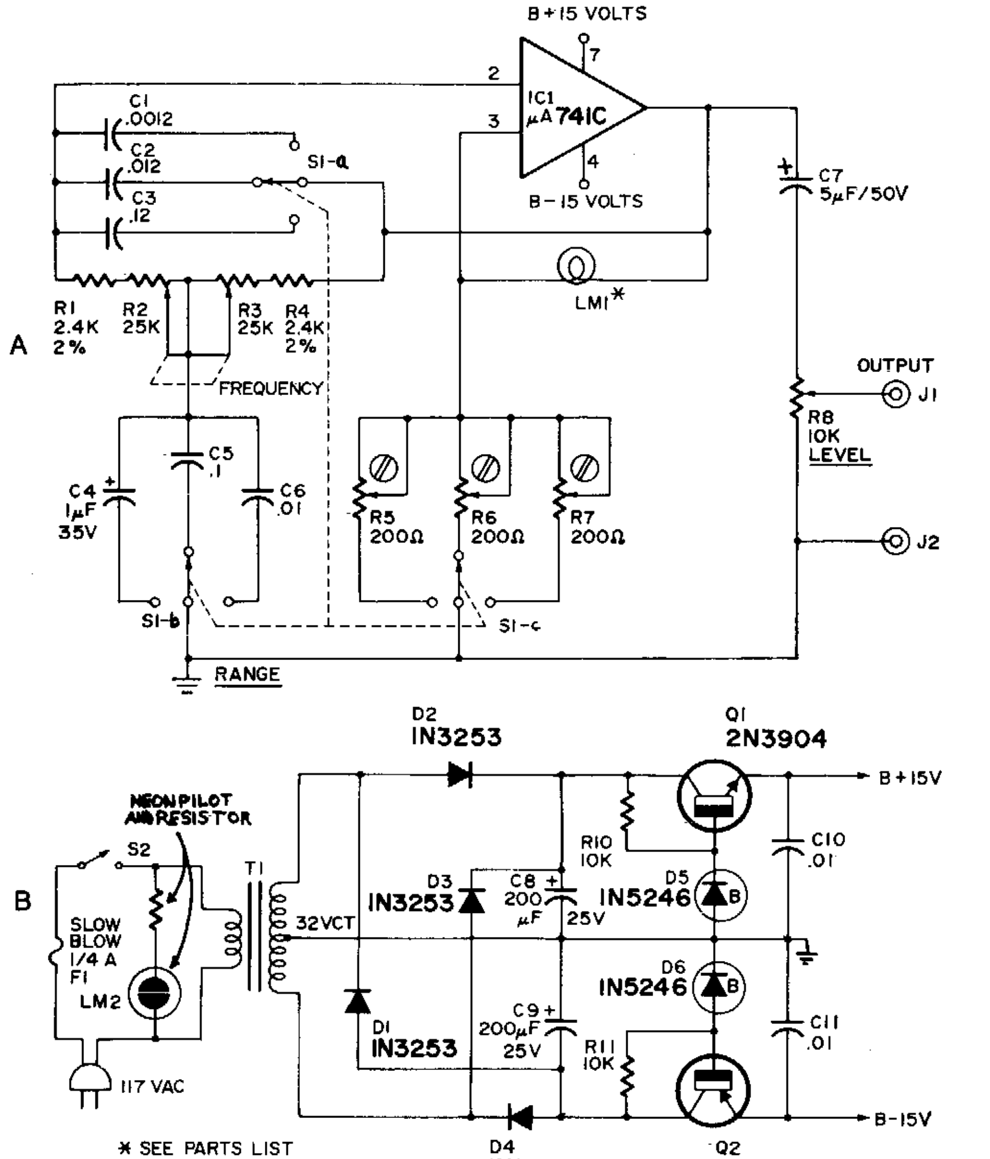
ELECTRONIC DESIGN

Fig. 88-9

Circuit Notes

Waveform purity at low frequencies for a Wien bridge oscillator is enhanced by diode limiting. Lamp L1 stabilizes the loop gain at higher frequencies while the limiting action of R2, CR1, and CR2 prevents clipping at low frequencies and increases the frequency adjustment range from about 3:1 to greater than 10:1.

ONE-IC AUDIO GENERATOR



RADIO-ELECTRONICS

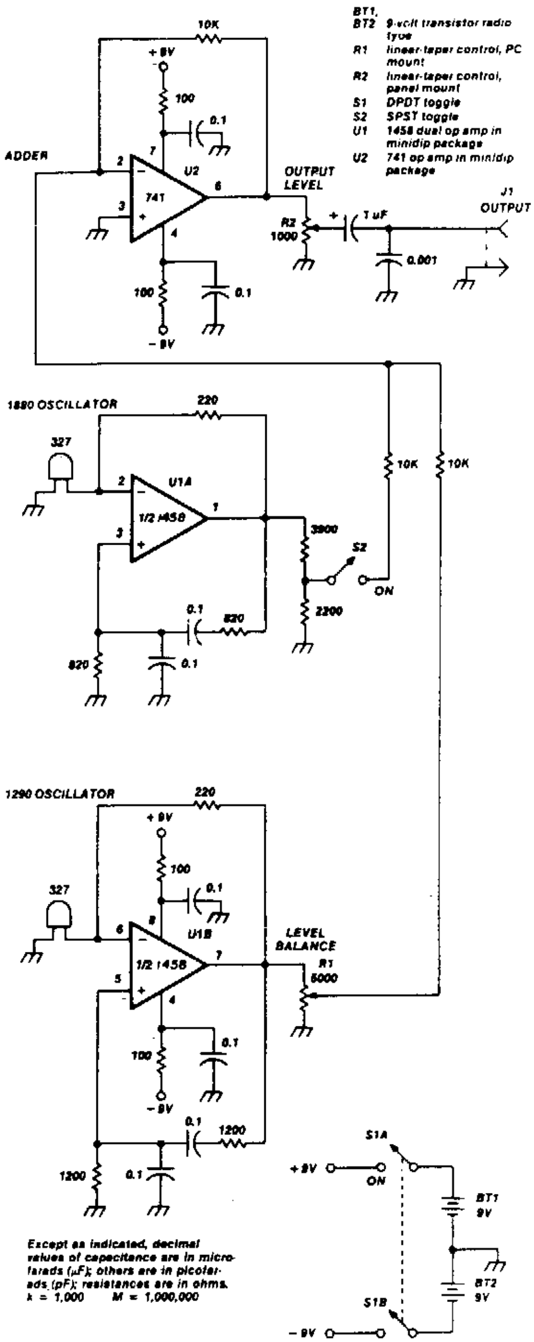
Fig. 88-10

Circuit Notes

This high-quality low-cost generator covers 20 Hz to 20 kHz in three bands with less than 1% distortion. LM1—10 V, 14 mA (344, 1869, 914) or 10 V, 10 mA (913, 367).

A = oscillator
B = power supply

SIMPLE TWO-TONE GENERATOR



HAM RADIO

Fig. 88-11

Circuit Notes

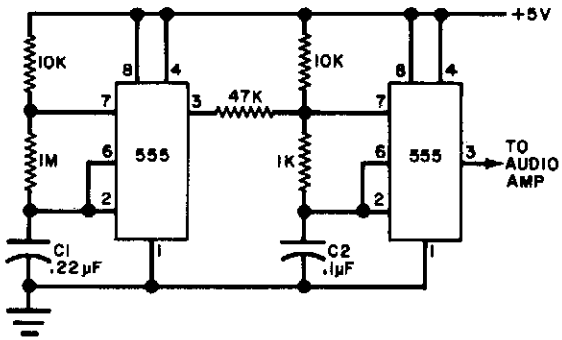
Two 741 operational amplifiers are used for the active element in this Wien bridge oscillator. (The 1458 is the dual version of the 741.) Frequencies of the two oscillators were chosen to fit standard component values. Other frequencies between 500 and 2000 Hz can be employed. They should not be harmonically related. The output level of U1A is set by a resistive divider, while the output of U1B is adjustable through R1. The output of the two oscillators is combined in U2, an op-amp adder with unity gain. The output from U2 can be adjusted using R2.

89

Sirens, Warblers and Wailers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

WARBLE GENERATOR

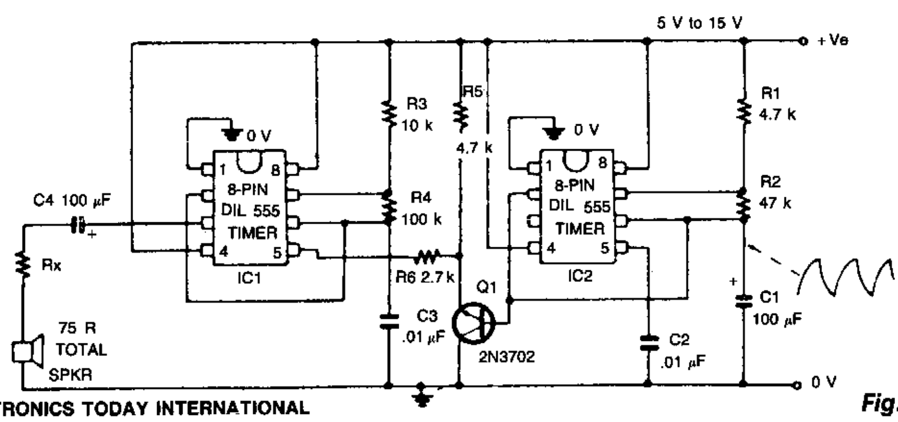


POPULAR ELECTRONICS *Fig. 89-1*

Circuit Notes

The circuit uses a pair of 555 timers or a single dual timer. Capacitor C1 controls the speed of the warble, while C2 determines the pitch. The values shown should produce quite a distinctive signal.

WAILING ALARM

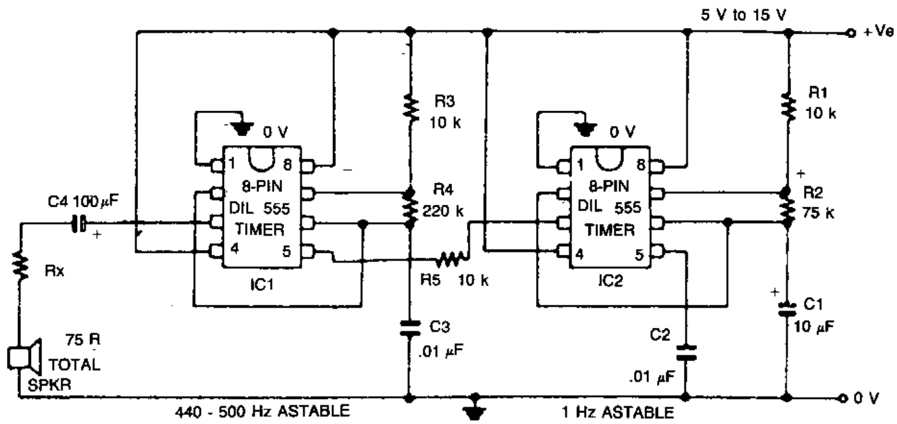


ELECTRONICS TODAY INTERNATIONAL *Fig. 89-2*

Circuit Notes

This circuit simulates the sound of an American police siren. IC2 is wired as a low frequency astable that has a cycling period of about 6 seconds. The slowly varying ramp waveform on C1 is fed to pnp emitter follower Q1, and is then used to frequency modulate alarm generator IC1 via R6. IC1 has a natural center frequency of about 800Hz. Circuit action is such that the alarm output signal starts at a low frequency, rises for 3 seconds to a high frequency, then falls over 3 seconds to a low frequency again, and so on ad infinitum.

WARBLE-TONE ALARM



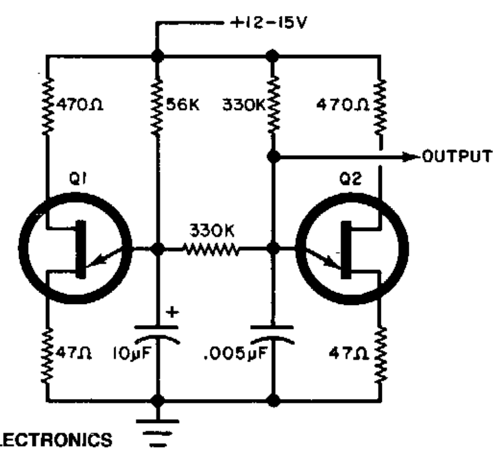
ELECTRONICS TODAY INTERNATIONAL

Fig. 89-3

Circuit Notes

The circuit generates a warble-tone alarm signal that simulates the sound of a British police siren. IC1 is wired as an alarm generator and IC2 is wired as a 1 Hz astable multivibrator. The output of IC2 is used to frequency modulate IC1 via R5. The action is such that the output frequency of IC1 alternates symmetrically between 500 Hz and 440 Hz, taking one sound to complete each alternating cycle.

WARBLING TONE GENERATOR



POPULAR ELECTRONICS

Fig. 89-4

Circuit Notes

The circuit use two unijunction transistors. The low-frequency sawtooth generated by Q1 modulates the high-frequency tone generated by Q2. The output should feed into a high-impedance amplifier. Q1 = Q2 = 2N4871.

MULTIFUNCTION SIREN SYSTEM

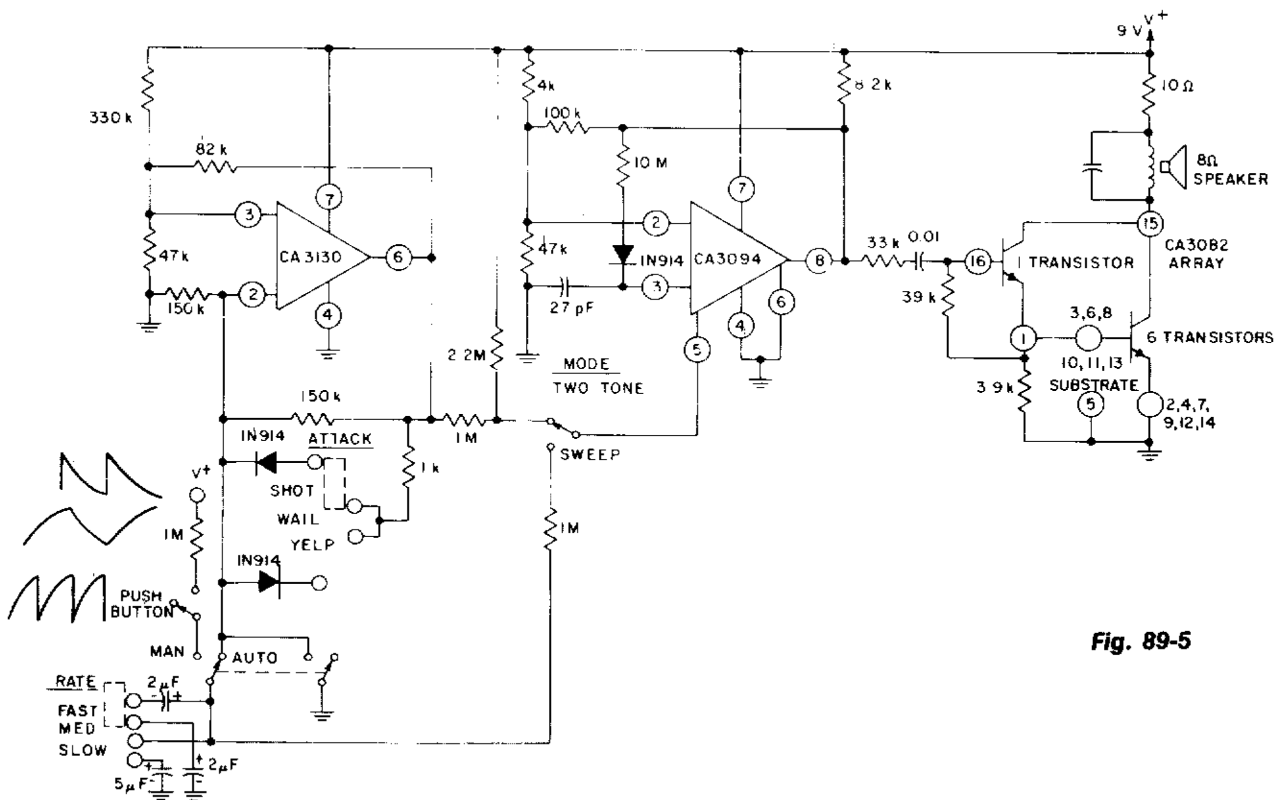


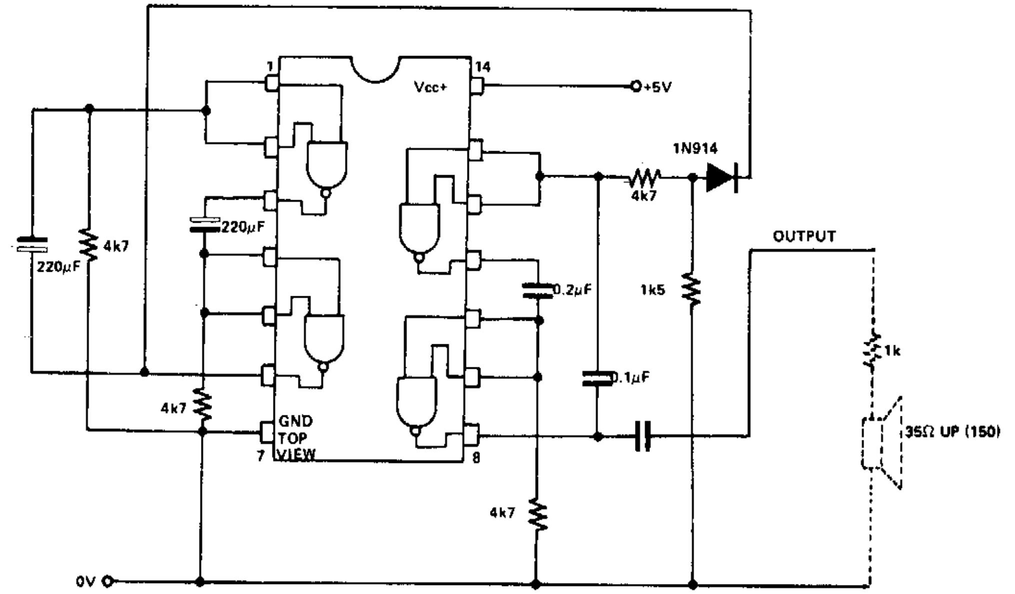
Fig. 89-5

GENERAL ELECTRIC/RCA

Circuit Notes

The circuit uses a CA3130 BiMOS op amp as a multivibrator to control the siren's rate. A CA3094 used as a VCO is followed by a CA3082 transistor array used to drive a speaker. A "Manual" or "Auto" mode switch allows the user to select either intermittent or continuous siren operation, respectively. In addition, three switches are available that control "Mode", "Attack", and "Rate".

7400 SIREN

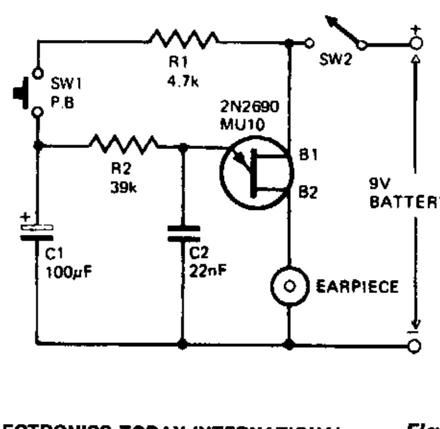


ELECTRONICS TODAY INTERNATIONAL *Fig. 89-6*

Circuit Notes

Two NAND gates are used for the oscillator, and two as the control. If the two-tone speed needs to be altered, the 220 μF capacitors can be changed (larger for slower operation). If the frequency of the oscillator is to be changed, the 0.2 and 0.1 μF capacitors can be varied and the value of R1 can be increased. To change frequency range between the two notes, alter the 1.5 k (1,500) resistor.

TOY SIREN

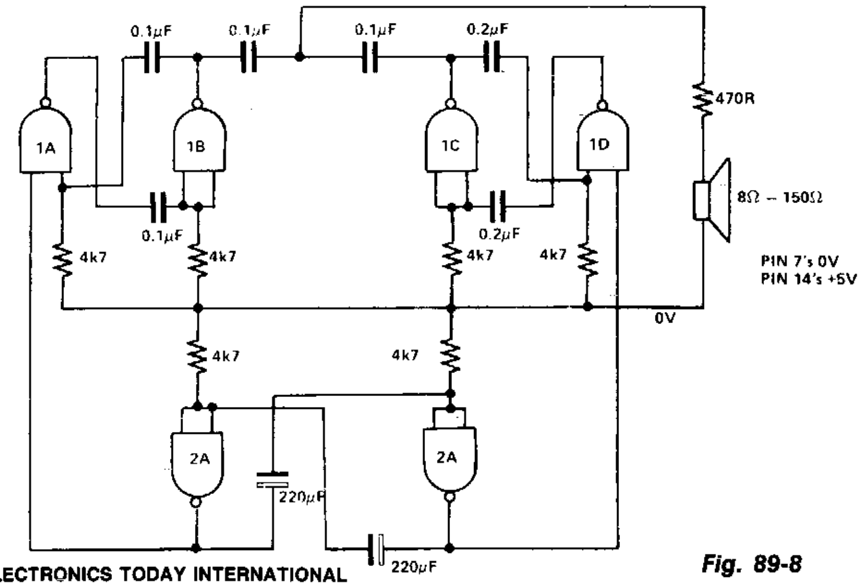


ELECTRONICS TODAY INTERNATIONAL *Fig. 89-7*

Circuit Notes

This circuit can be built small enough to fit inside a toy. The circuit consists of a relaxation oscillator utilizing one unijunction transistor (2N2646, MU10, TIS43). R2 and C2 determine the frequency of the tone. Pushing the button, SW1 charges up the capacitor and the potential at the junction of R2 and C2 rises, causing an upswing in the frequency of oscillation. On releasing the pushbutton the charge on C2 will drop slowly with a proportional reduction in the frequency of oscillation. Manual operation of the button at intervals of approximately 2 seconds will produce a siren sound.

SIREN USES TTL GATES



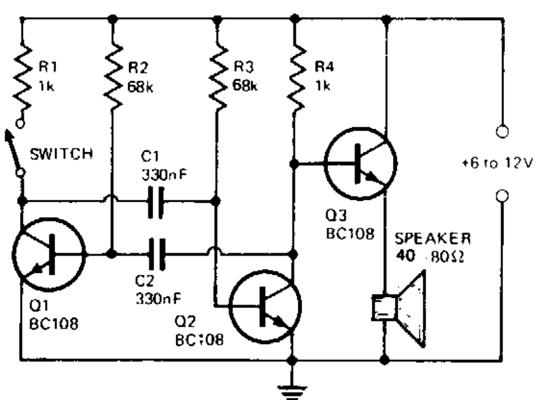
ELECTRONICS TODAY INTERNATIONAL

Fig. 89-8

Circuit Notes

The siren consists of two oscillators which generate the tones. A third oscillator is used to switch the others on and off alternately, giving the two-tone effect. By changing the capacitor values different tones can be produced.

ELECTRONIC SHIP SIREN



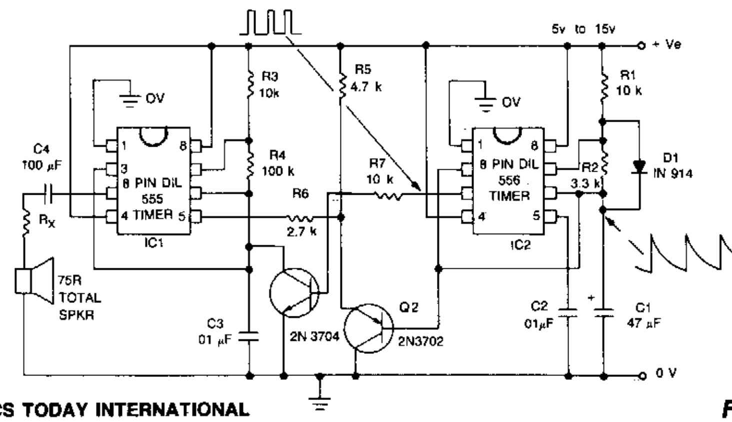
ELECTRONICS TODAY INTERNATIONAL

Fig. 89-9

Circuit Notes

The circuit consists of a multivibrator (Q1 & Q2), and a low power output stage Q3. The speaker should have an impedance in the region of 40 to 80 ohms. To use a low impedance speaker, connect an output transformer from the emitter of Q3 to ground. C1 and C2 determine the pitch of the siren and the values specified will provide a tone of about 300 Hz. Quiescent current is negligible. The output at the collector of Q2 can also be fed into an amplifier input via a 1 μF electrolytic, in series with a 12 k resistor.

SIREN ALARM SIMULATES STAR TREK RED ALERT



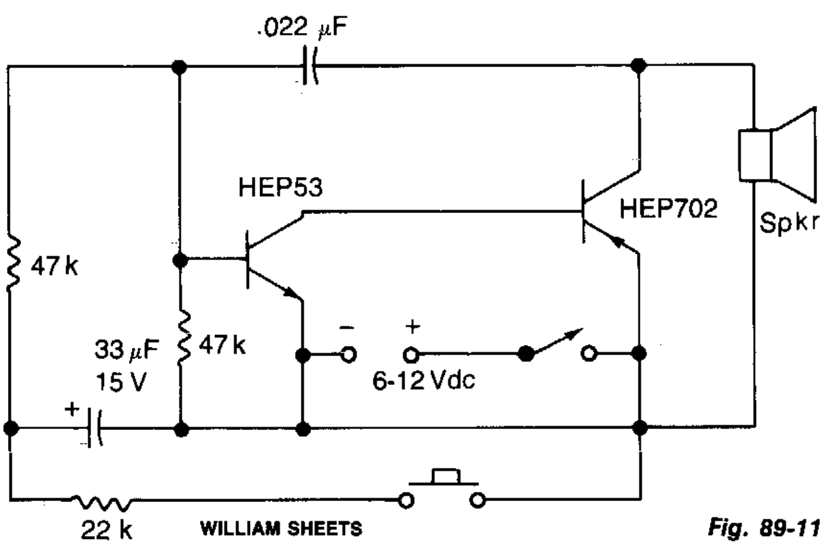
ELECTRONICS TODAY INTERNATIONAL

Fig. 89-10

Circuit Notes

The signal starts at a low frequency, rises for about 1.15 seconds to a high frequency, ceases for about 0.35 seconds, then starts rising again from a low frequency, and so on ad infinitum.

YELP OSCILLATOR



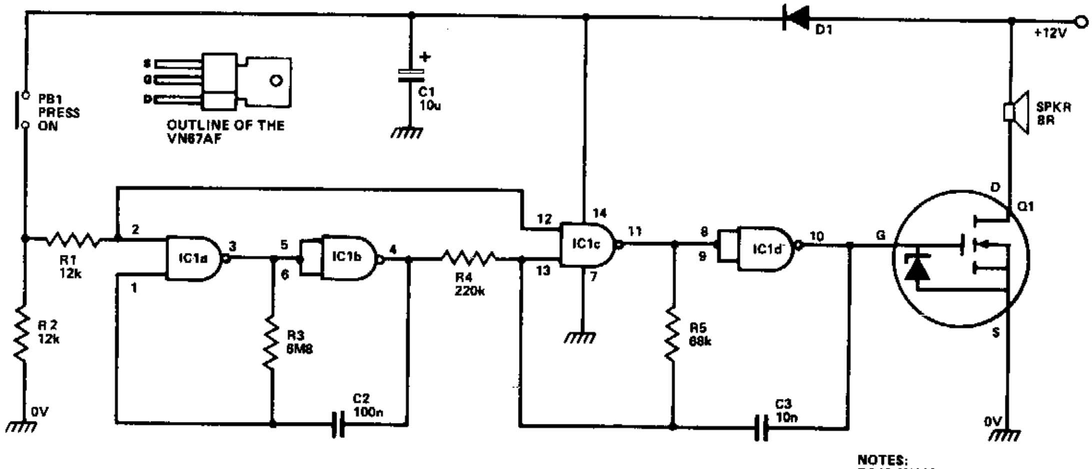
WILLIAM SHEETS

Fig. 89-11

Circuit Notes

Close the pushbutton switch and the circuit starts the siren up-shifting to a higher frequency. Release it and the tone slides down until S2 is closed again. Tone quality is adjusted by changing the 0.022 μF capacitor.

HIGH POWER SIREN



NOTES:
D1 IS 1N4001
Q1 IS VN67AF
IC1 IS CD4011B

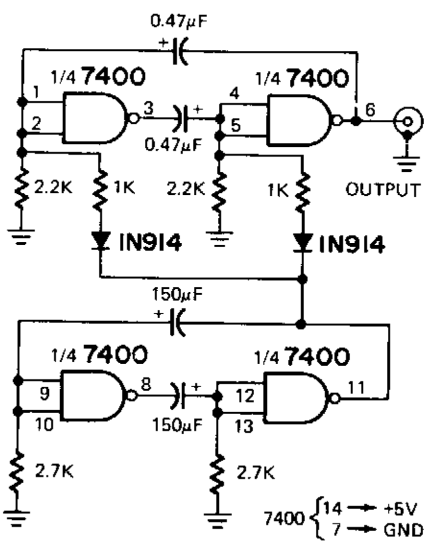
ELECTRONICS TODAY INTERNATIONAL

Fig. 89-12

Circuit Notes

IC1a and IC1b are wired as a slow astable multivibrator and IC1c-IC1d are wired as a fast astable. Both are "gated" types, which can be turned on and off via PB1. The output of the slow astable modulates the frequency of the fast astable, and the output of the fast astable is fed to the external speaker via the Q1 VMOS power FET amplifier stage.

"HEE-HAW" TWO-TONE SIREN



RADIO-ELECTRONICS

Fig. 89-13

Circuit Notes

The circuit uses two gates of a 7400 IC cross-connected to form an astable multivibrator driven by the 1-pulse per second output of the digital clock IC. The hee-haw circuit has a low frequency astable modulator added to make a self-contained European-type siren. Tone and rate can be varied as desired by changing capacitor values. If the tone is too harsh, a simple R-C filter will remove the harmonic content—the multivibrator output is almost a square wave. With the resistor values shown, no start-up problems occur; but if the 2.2 k or 2.7 k resistors are changed too much, latch-up can be a problem.

VARYING-FREQUENCY WARNING ALARM

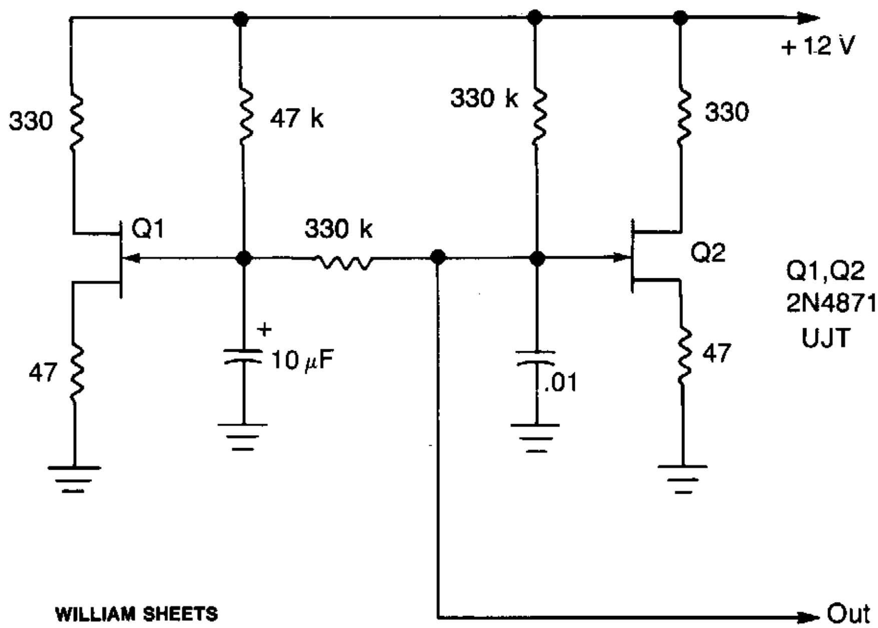


Fig. 89-14

Circuit Notes

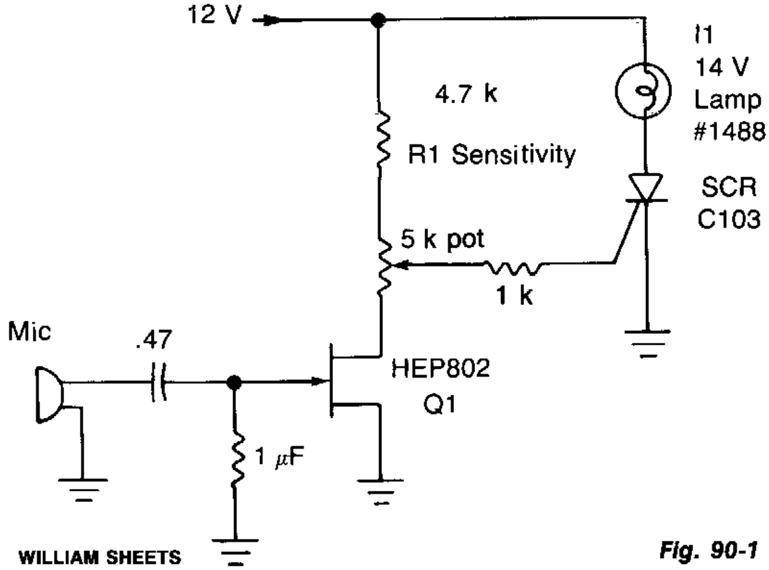
The output frequency changes continuously. Low frequency oscillator (Q1) modulates high frequency oscillator Q2 and its associated timing capacitor.

90

Sound (Audio) Operated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SOUND-ACTIVATED SWITCH



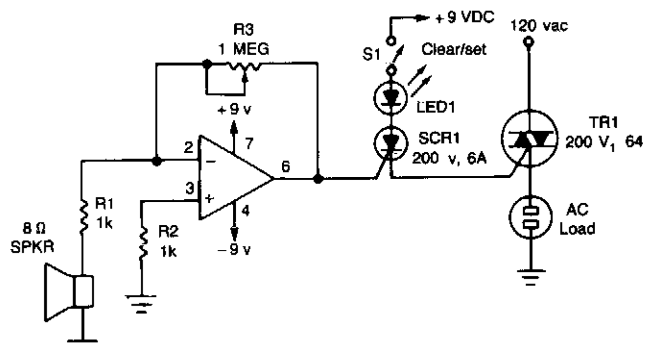
WILLIAM SHEETS

Fig. 90-1

Circuit Notes

The audio from Mic is amplified by Q1. Peaks of signal (adjusted by R1) greater than about 0.7 volts trigger the SCR and light lamp I1.

SOUND-ACTIVATED ac SWITCH



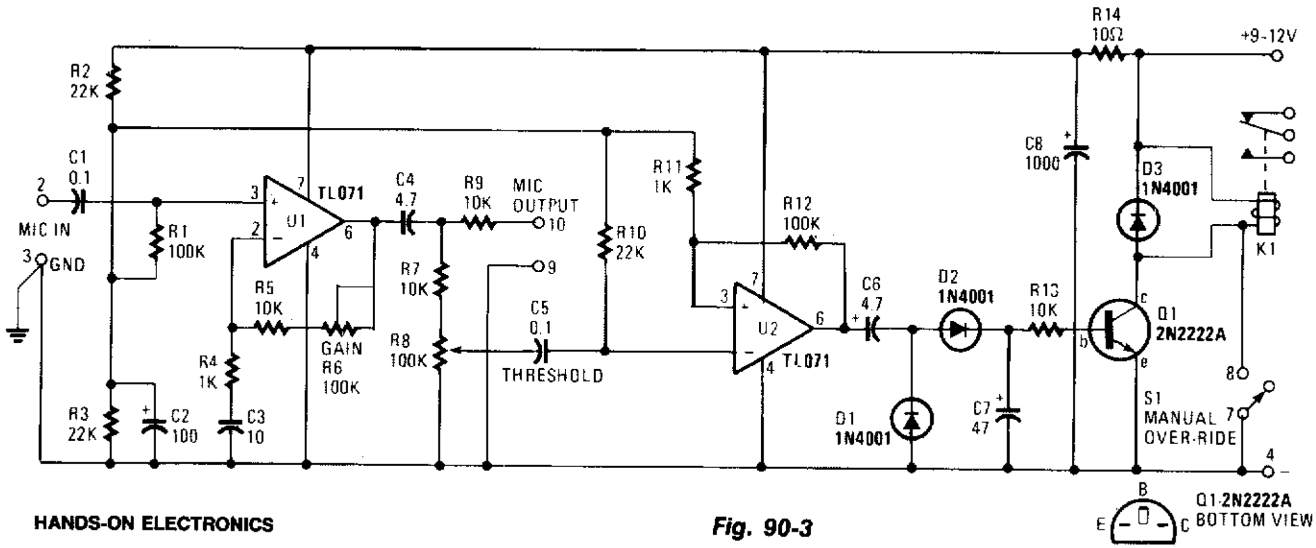
RADIO-ELECTRONICS

Fig. 90-2

Circuit Notes

The circuit uses a 741 op amp operating as an inverting amplifier to amplify the voltage produced by an 8-ohm speaker used to detect any sounds. The feedback resistor R3, a 1-megohm potentiometer used to vary the gain of the amplifier determines the sensitivity of the circuit. When S1 is closed in the (SET) position and a sound is applied to the speaker, SCR1 is turned on. It will remain in conduction until the anode voltage is removed by opening S1, putting it in its RESET position. (Once an SCR is turned on, the gate or trigger has no control over the circuit.) As long as the SCR conducts, the Triac, TR1, will remain on and supply voltage to the load.

VOX BOX



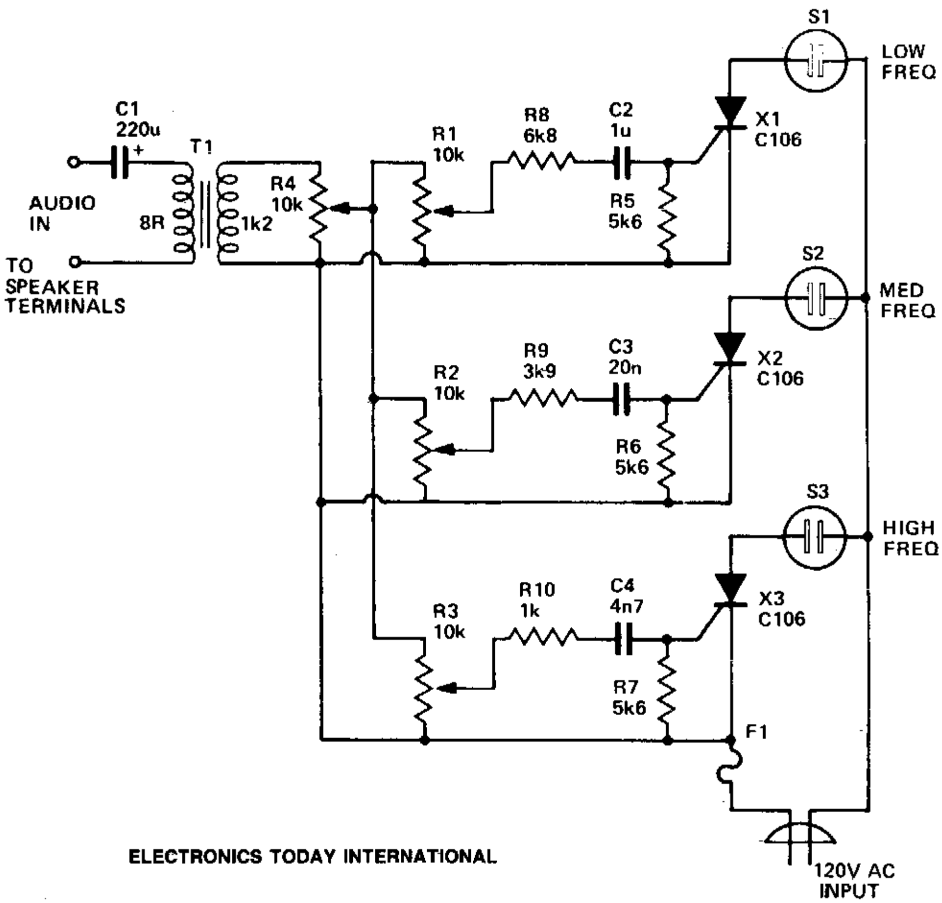
HANDS-ON ELECTRONICS

Fig. 90-3

Circuit Notes

The electronic circuit in the VOX Box consists of three parts: a microphone preamplifier, a Schmitt trigger, and a relay driver. Input signals (MIC INPUT terminals) to the microphone preamplifier (U1) are amplified and fed to a THRESHOLD control (R8). When the preselected threshold voltage level is exceeded, the output of the Schmitt trigger (U2) immediately goes high. The signal from U2 is rectified and the voltage developed across C7 turns on the relay energizer transistor (Q1). That transistor action passes pull-down current through the coil of relay K1. The changing of the relay SPDT contacts can be used to either make or break an external ac or dc circuit.

COLOR ORGAN



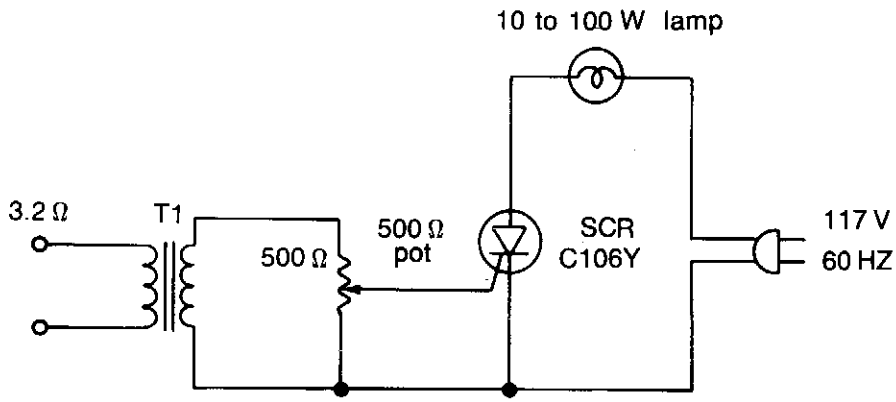
ELECTRONICS TODAY INTERNATIONAL

Fig. 90-4

Circuit Notes

Three lights are controlled by the three channels. One light will pulse in response to the bass, another illuminates with mid-range sounds, and the last lights for high notes. Four level controls allow adjustment of overall light level and each channel individually. Up to 200 watts per channel can be handled.

BASIC COLOR ORGAN



WILLIAM SHEETS

Fig. 90-5

Circuit Notes

Transformer T1 can be any matching transistor type in the range of 500/500 to 2500/2500 ohms. No connections from the SCR or its components are connected to ground. For safety's sake, keep the 117-V line voltage from the amplifier connections—that is the reason for using T1. To adjust, set potentiometer R1 ''off'' and adjust the amplifier volume control for a normal listening level. Then adjust the potentiometer until the lamp starts to throb in step with the beat.

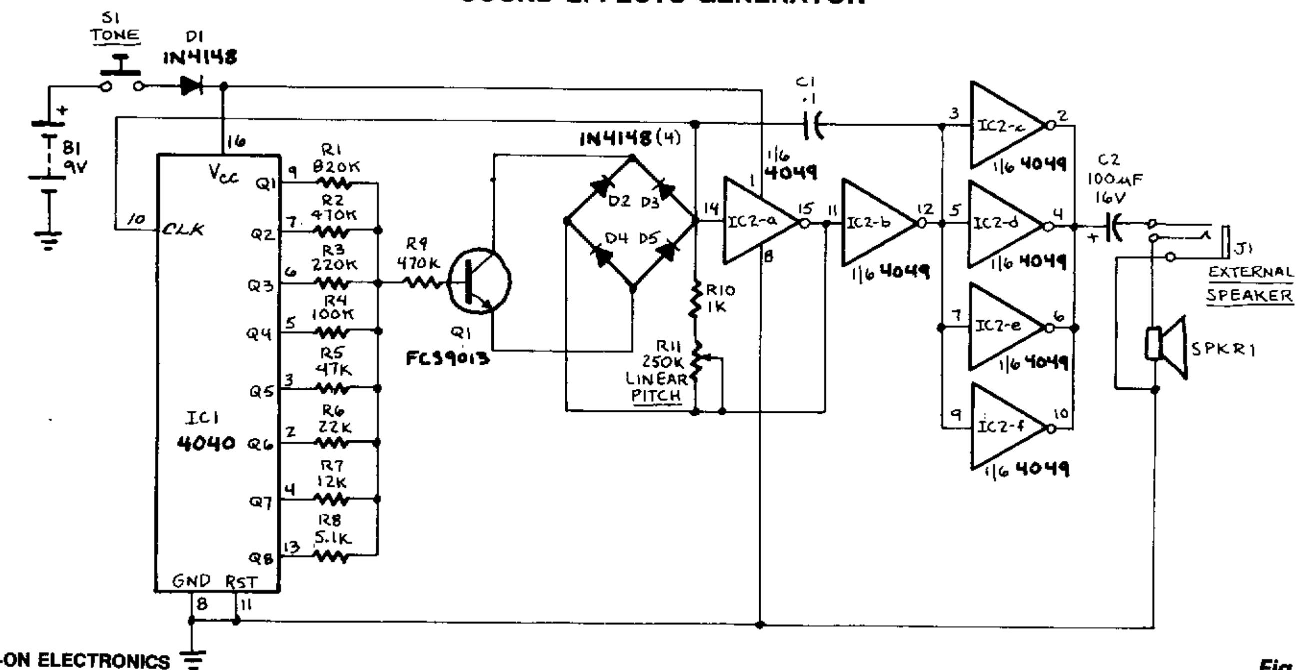
91

Sound Effect Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Sound Effects Generator
Electronic Bongos
Train Chuffer
Bird Chirp
Steam Locomotive Whistle
WAA-WAA Circuit
Unusual Fuzz
Autodrum
Twang-Twang Circuit
Steam Train/Prop Plane
Funk Box

SOUND EFFECTS GENERATOR



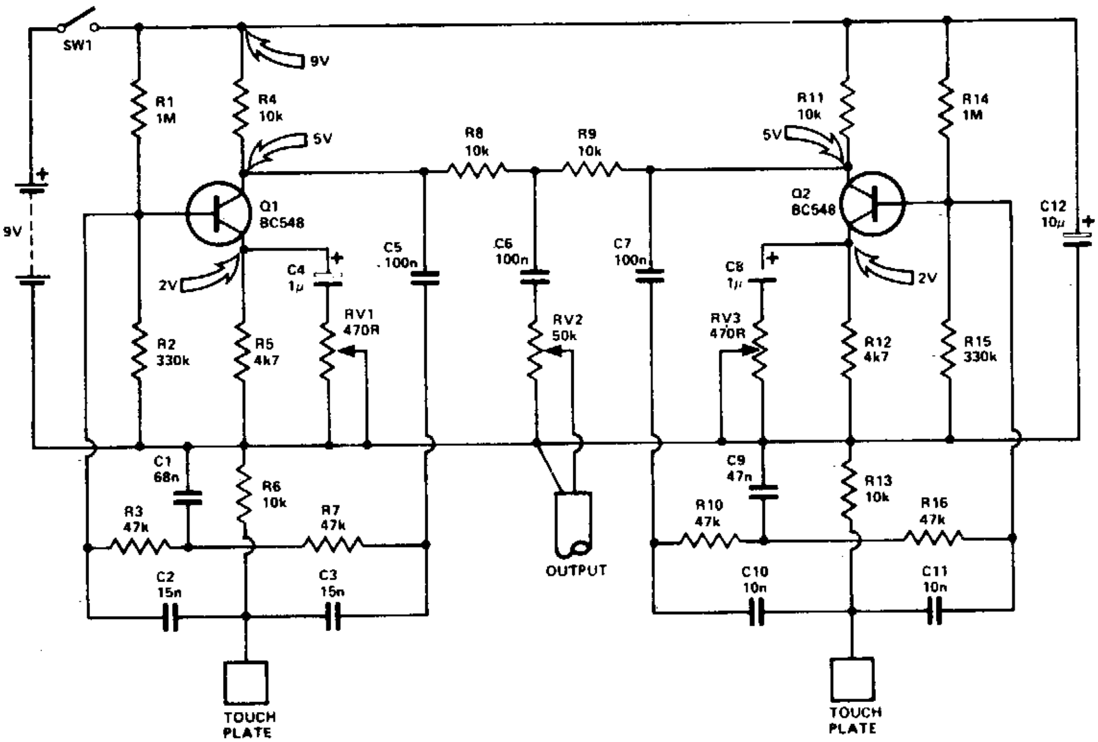
HANDS-ON ELECTRONICS

Fig. 91-1

Circuit Notes

This circuit can generate a European police-car siren, bird noises, spaceship sounds, etc. In addition, it can be put to use as a doorbell, an alarm, etc. The circuit consists of four parts: a binary counter, a D/A converter, a VCO, and an audio output amplifier. The initial frequency of oscillation is determined by potentiometer R11. The VCO first oscillates at a relatively low frequency, and gradually picks up speed as the control voltage supplied by the D/A converter increases. The D/A converter is the group of resistors R1-R8. When none of IC1's outputs is active, little current will flow into the base of Q1, so the VCO's control voltage will be low. As more and more counter outputs become active, base current increases, and thereby so does the VCO's frequency of oscillation. The VCO itself is composed of IC2-a, IC2-b, Q1, and the timing network comprising D1-D4, C1, R10, and R11. The buffer amplifier is made up of the four remaining gates from IC2, all wired in parallel.

ELECTRONIC BONGOS

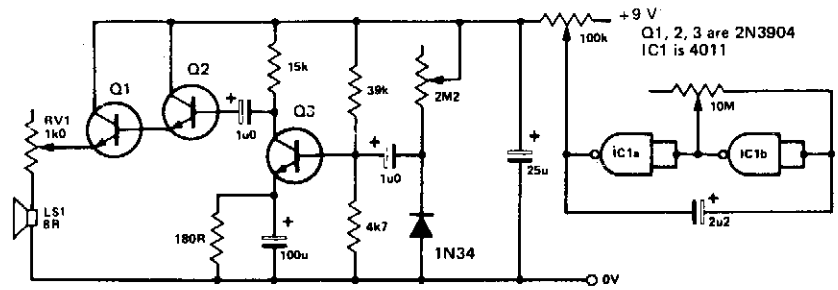


ELECTRONICS TODAY INTERNATIONAL

Circuit Notes

This circuit consists of twin-T sine-wave oscillators. Each oscillator has a filter in the feedback loop. If the loop gain is greater than unity, the circuit will oscillate. Gain is adjusted to be just less than unity. Touching the touch plate starts the oscillator, but the moment your finger is removed from the touch plate the oscillations will die away. The rate of decay is a function of circuit gain and controlled by RV1 (and RV3).

TRAIN CHUFFER



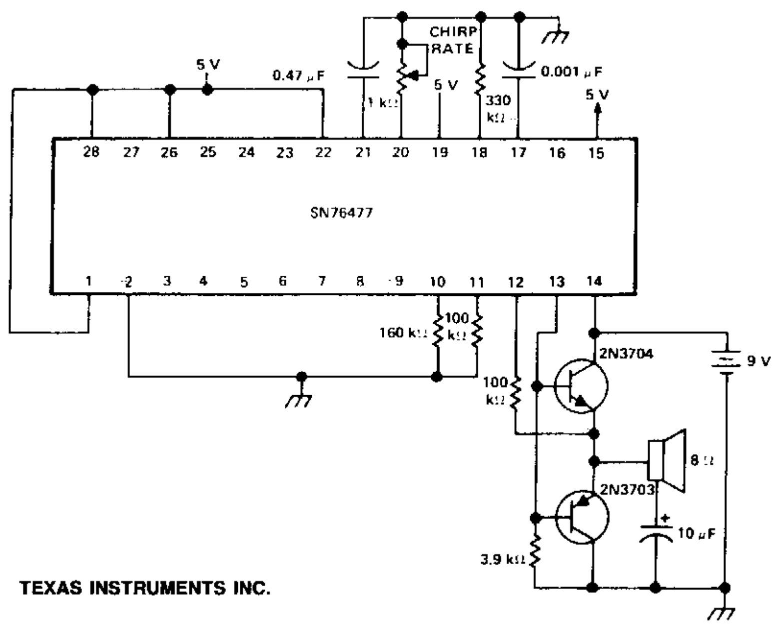
ELECTRONICS TODAY INTERNATIONAL

Fig. 91-3

Circuit Notes

The circuit consists of a white noise generator which only switches on with the high part of the square wave output from the clock circuit. The frequency of the clock is adjusted with the 10 M pot and the output voltage of the clock is adjusted by the 100 k pot (rate and volume of chuff respectively). The 2M2 pot controls the amount of noise produced and the 1 k pot on the speaker controls the pitch of the average noise.

BIRD CHIRP



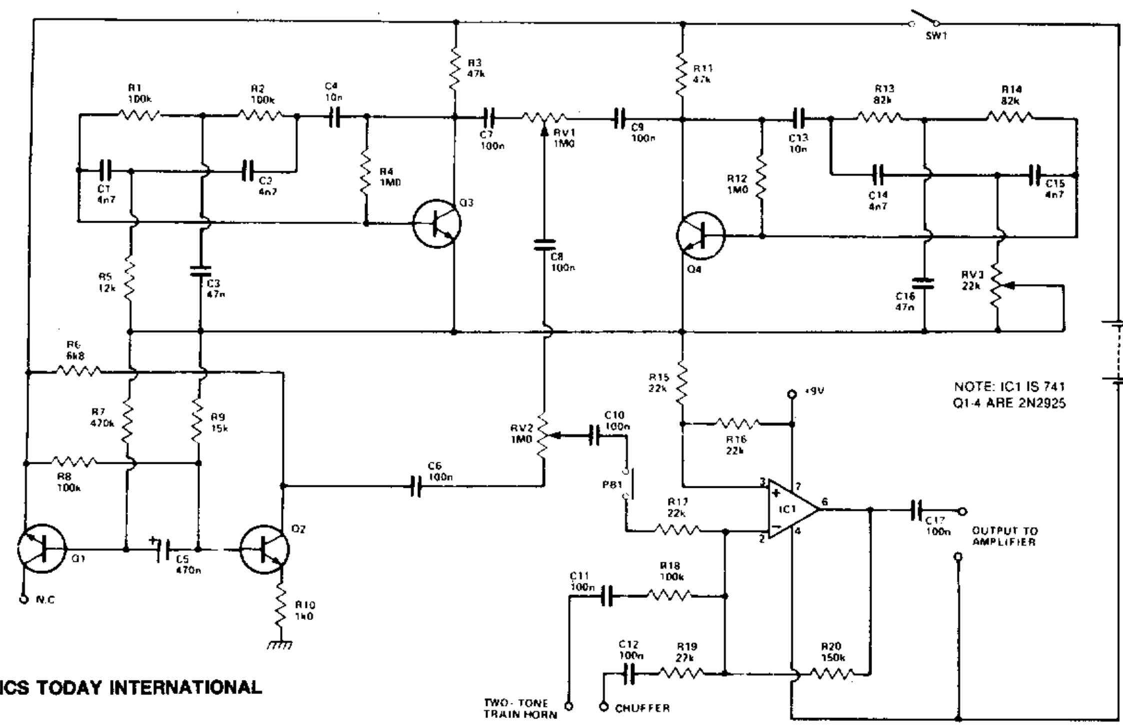
TEXAS INSTRUMENTS INC.

Fig. 91-4

Circuit Notes

For a barking dog, the capacitor at pin 17 is changed to 15 pF to increase the frequency of the VCO.

STEAM LOCOMOTIVE WHISTLE



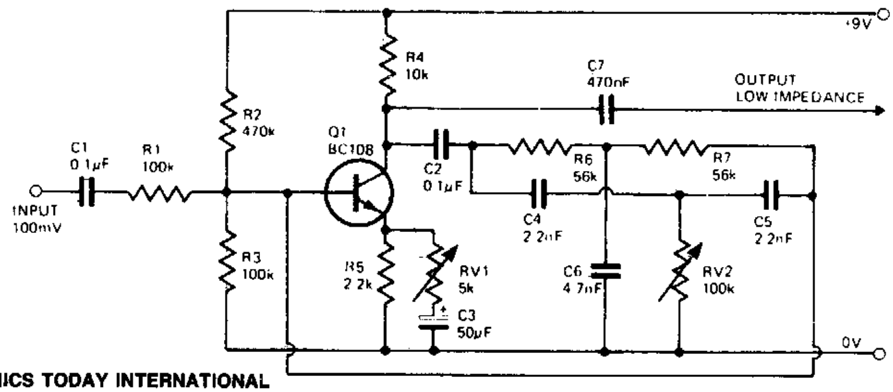
ELECTRONICS TODAY INTERNATIONAL

Fig. 91-5

Circuit Notes

The waveform of a steam whistle is a complex combination of white noise and an audio frequency oscillation. The noise generator is a transistor (Q1) biased into zener mode. The audio frequency oscillation is a straightforward mixture of two similar (but not identical) sine waves, which after their addition produce a more complex waveshape. The sine wave generators are twin-t oscillators. Preset RV1 mixes the two sine waves so that an appropriate waveform is obtained. RV2 mixes this waveform with the white noise. Adjustment of all three presets will result in the required sound. Integrated circuit IC1 is an operational amplifier used as a simple mixer/amplifier which combines the steam whistle, chuffer, (generated elsewhere) and two-tone horn sounds into one, suitable for amplification by an external amplifier.

WAA-WAA CIRCUIT



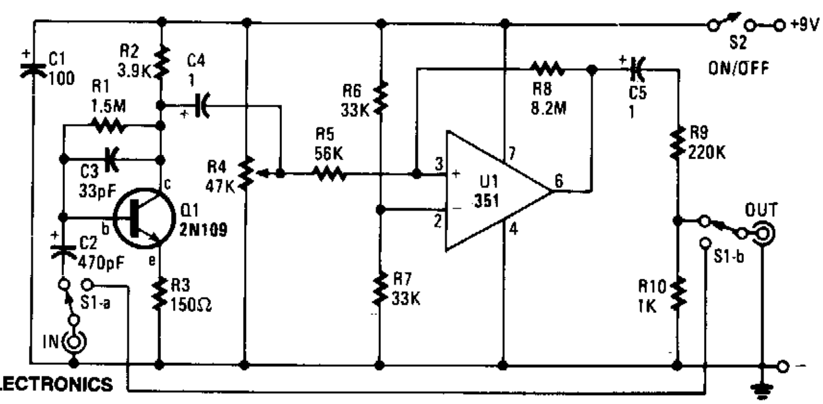
ELECTRONICS TODAY INTERNATIONAL

Fig. 91-6

Circuit Notes

The waa-waa effect is achieved as certain frequencies are amplified more than others. A phase shift RC oscillator makes up the basic circuit. Negative feedback is obtained by feeding part of the signal back to the base. When adjusting initially, RV1 is turned to minimum. RV2 is adjusted to a point at which an audible whistle appears indicating oscillation. RV1 is then adjusted till the oscillation just disappears. It should be possible to set RV2 to any value without any oscillation, this should also be achieved with the minimum possible value of RV1.

UNUSUAL FUZZ



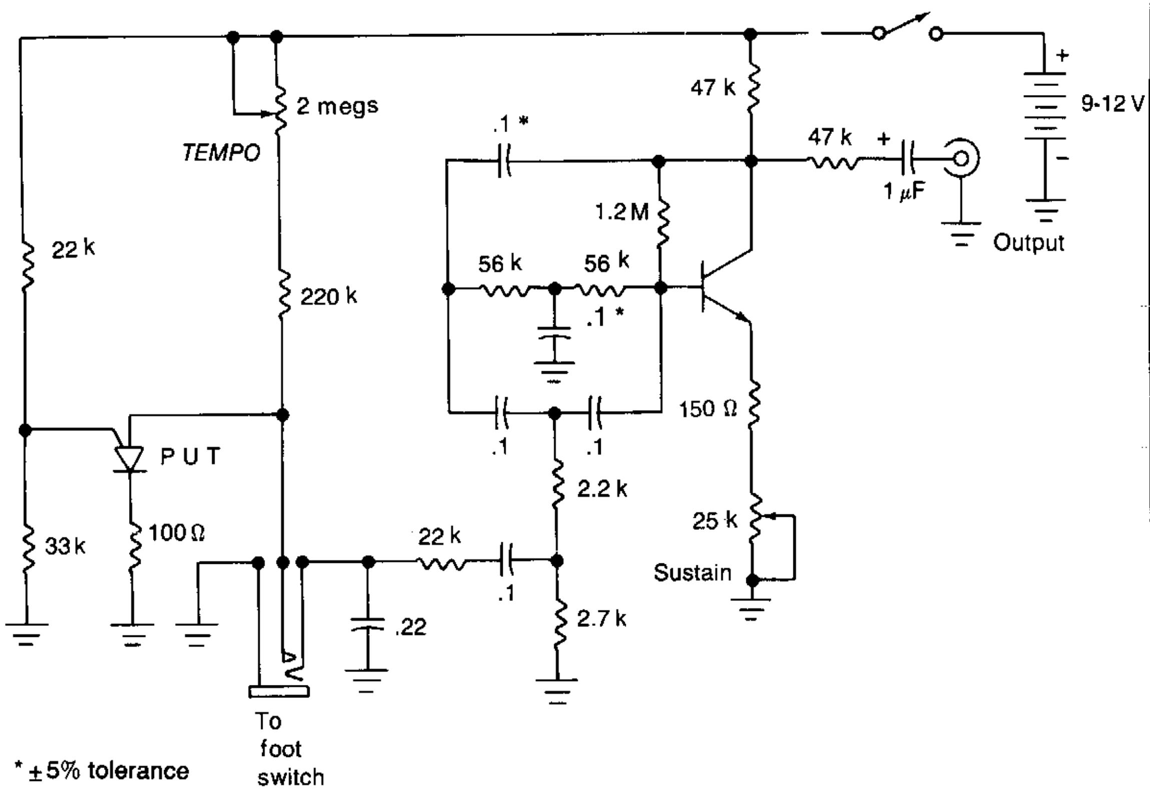
HANDS-ON ELECTRONICS

Fig. 91-7

Circuit Notes

It seems that guitar fuzz boxes have been around since the beginning of rock, and have seen little improvement over the years. This one is somewhat different because rather than simply distorting the sound, it also pulses in step with the peaks of the waveform from the pickup because of the Schmitt trigger op amp circuit. Capacitor C2 requires some explanation. It should normally be a 1- or 2-μF electrolytic capacitor. However, we show the value as 470 pF because it's recommended as an experimental value giving far out effects.

AUTODRUM



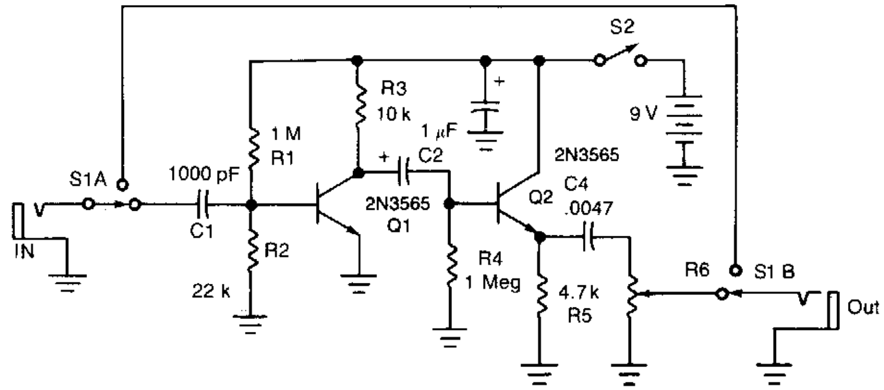
WILLIAM SHEETS

Fig. 91-8

Circuit Notes

This unit generates a drum-like damped oscillation that sounds best when fed into a higher power amplifier. The beat rate may be determined by operating a foot pedal in much the same manner as for a real drum, or by means of an internal oscillator, the speed of which may be preset.

TWANG-TWANG CIRCUIT



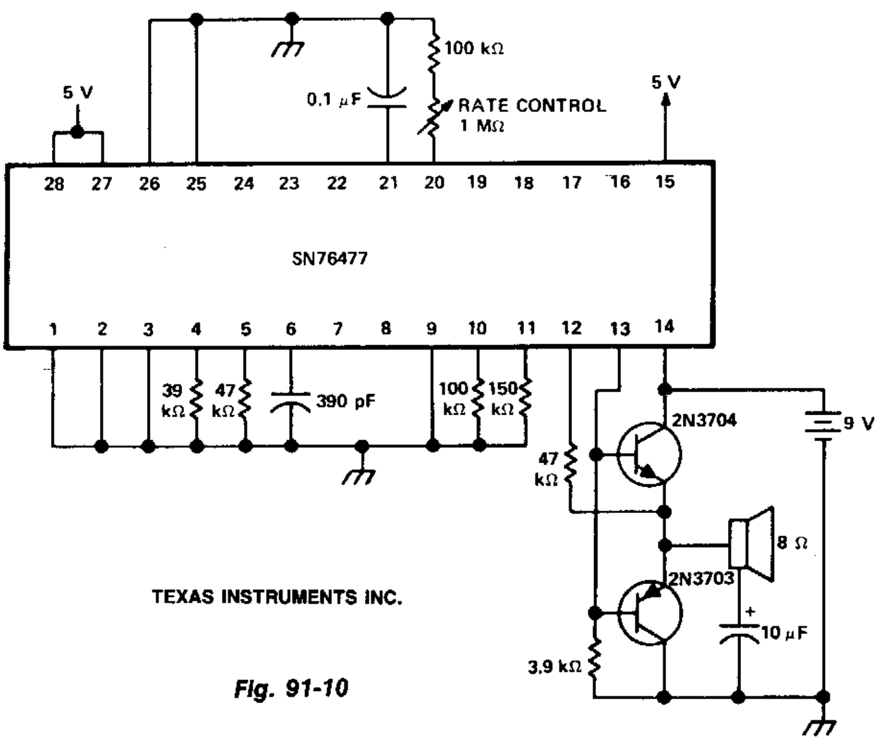
WILLIAM SHEETS

Fig. 91-9

Circuit Notes

Twang is a guitar sound that more or less approximates a banjo or mandolin. The circuit produces unusual sounds from an ordinary electric guitar by cutting the bass, severely distorting the midband and highs, and then amplifying the distortion. S1 cuts the effect in and out, S2 turns the unit on and off.

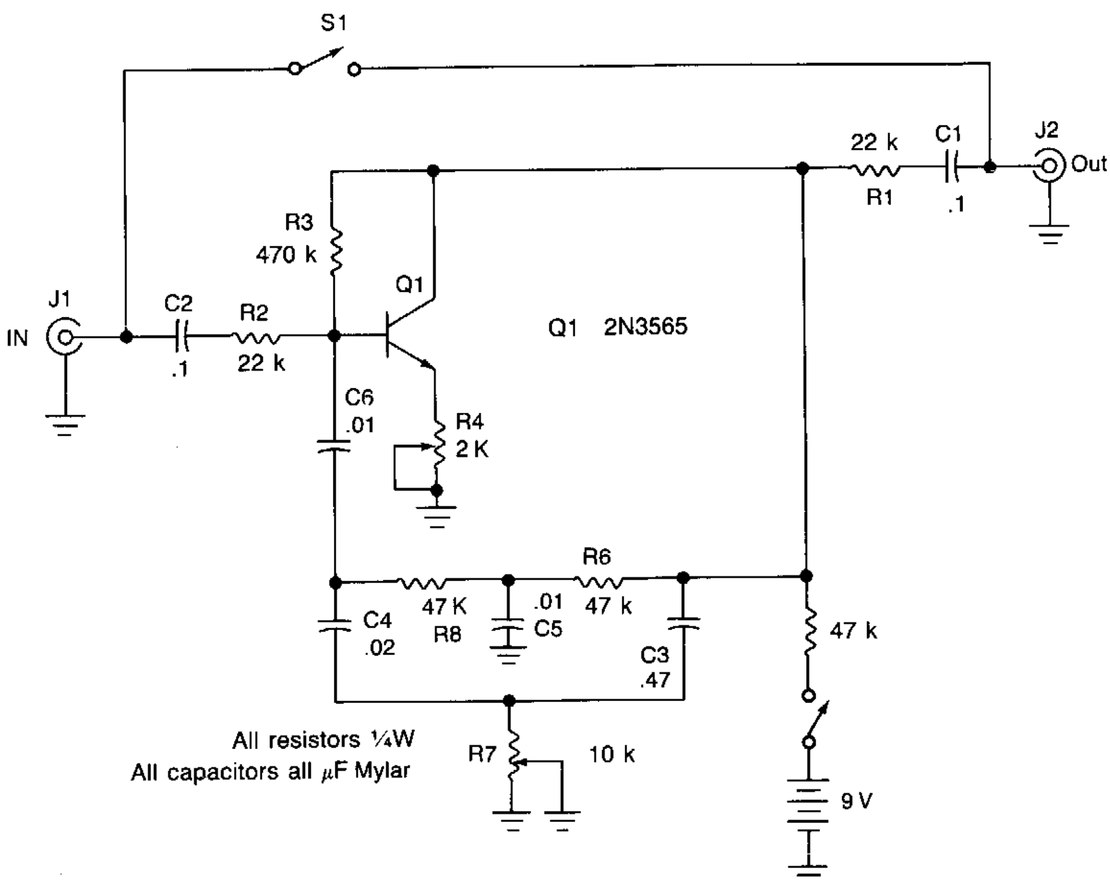
STEAM TRAIN/PROP PLANE



TEXAS INSTRUMENTS INC.

Fig. 91-10

FUNK BOX



WILLIAM SHEETS

Fig. 91-11

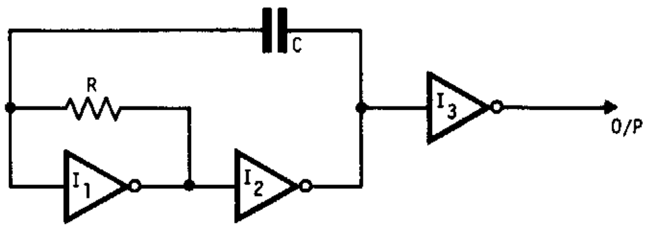
Circuit Notes

Adjusting potentiometer R7 adds extra twang from way down low to way up high. To set the unit, adjust potentiometer R4 until you hear a whistle (oscillation); then back off R4 until the oscillation just ceases. The effect can be varied from bass to treble by R7.

92
Square-Wave Generators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LOW FREQUENCY TTL OSCILLATOR

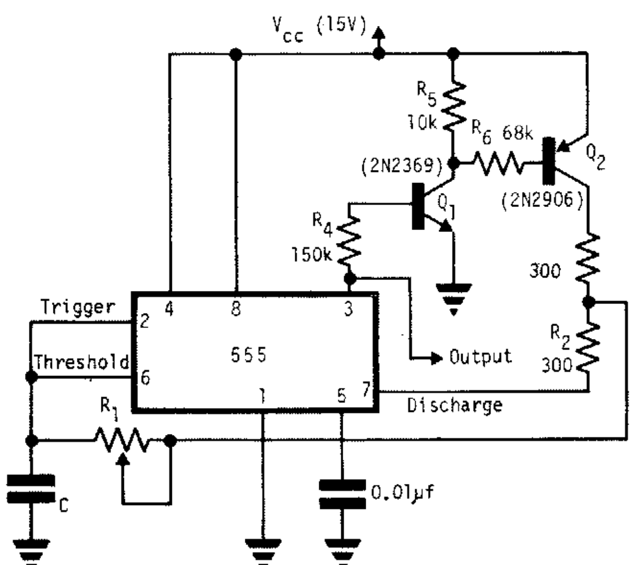


ELECTRONIC ENGINEERING *Fig. 92-1*

Circuit Notes

This oscillator uses standard inverters, one resistor and one capacitor, and has no minimum operating frequency. R and C must be chosen such that currents into the gates are below recommended operating limits and that leakage current into the gates and into C are small in comparison with the current in R also the output should be buffered (I3) to prevent variations in load affecting frequency. This circuit may also be used to square up slowly changing logic levels by use of multi input gates (NANDS, NORS Etc).

SQUARE-WAVE GENERATOR USING A 555 TIMER

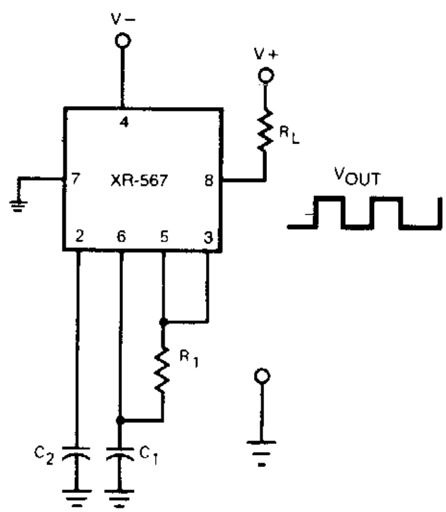


ELECTRONIC ENGINEERING *Fig. 92-2*

Circuit Notes

A single timing resistor ensures that the output is a square (50% duty cycle) wave at all frequency settings.

OSCILLATOR WITH FREQUENCY DOUBLED OUTPUT



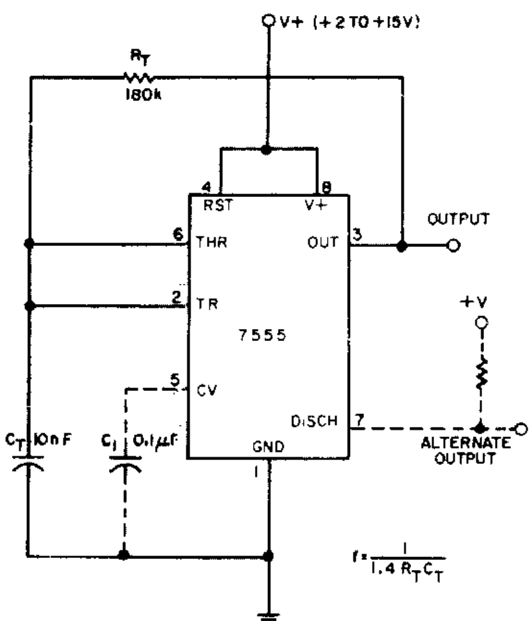
EXAR

Fig. 92-3

Circuit Notes

The current-controlled oscillator frequency can be doubled by applying a portion of the square-wave output at pin 5 back to the input at pin 3, as shown. In this manner, the quadrature detector functions as a frequency doubler and produces an output of 2 f_0 at pin 8.

CMOS 555 ASTABLE GENERATES TRUE RAIL-TO-RAIL SQUARE WAVES



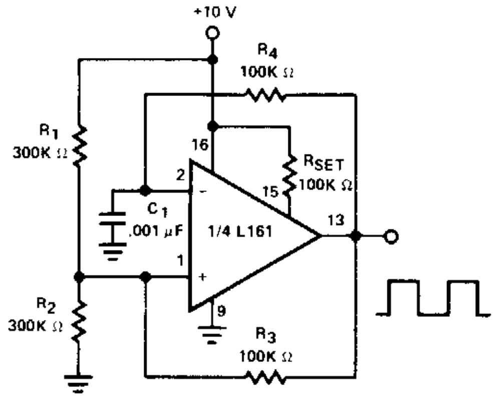
ELECTRONIC DESIGN

Fig. 92-4

Circuit Notes

A CMOS timer generates true square waves because, unlike the bipolar 555, its output swings from rail to rail. The component values shown give a frequency of about 400 Hz.

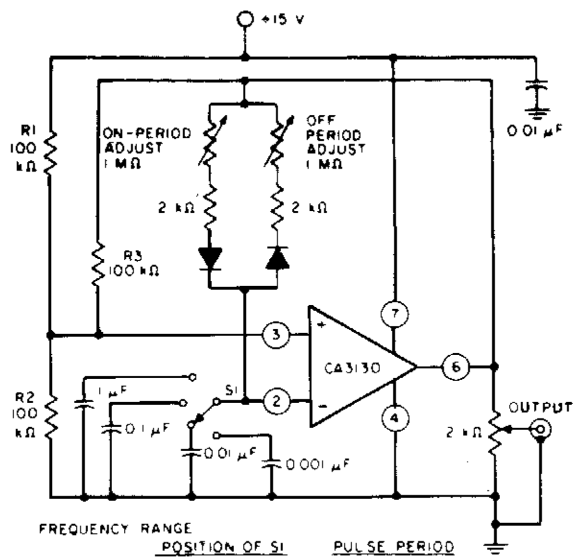
SQUARE-WAVE OSCILLATOR



SILICONIX

Fig. 92-5

ASTABLE MULTIVIBRATOR



FREQUENCY RANGE

POSITION OF S1	PULSE PERIOD
0.001 µF	4 µs TO 1 ms
0.01 µF	40 µs TO 10 ms
0.1 µF	0.4 ms TO 100 ms
1 µF	4 ms TO 1 s

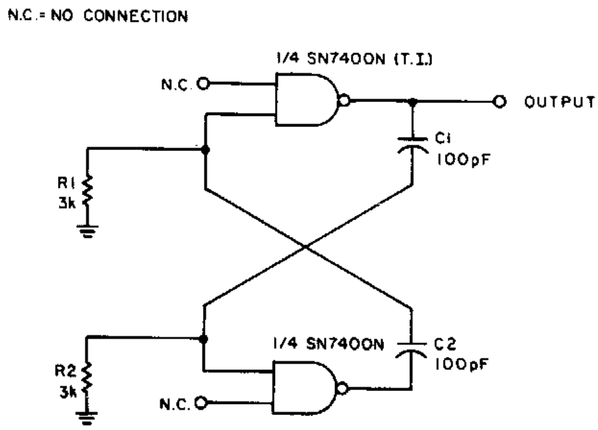
GENERAL ELECTRIC/RCA

Fig. 92-6

Circuit Notes

The circuit with independent control of "ON" and "OFF" periods uses the CA3130 BiMOS op amp for filters, oscillators, and long-duration timers. With input current at 50 pA, oscillators can utilize large-resistor/small-capacitor combinations without loading effects.

TWO-MHz SQUARE-WAVE GENERATOR USES TWO TTL GATES



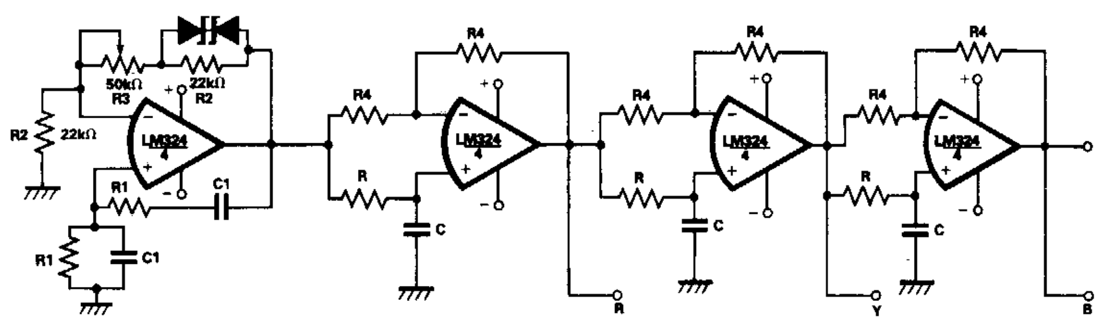
ELECTRONIC DESIGN

Fig. 92-7

Circuit Notes

With the values shown the circuit generates a 2-MHz symmetrical square wave. Changing capacitors C1 and C2 to 0.01 μF results in a frequency of 500 Hz. For the particular integrated circuits and power supply voltages (5.0 V), the reliable operating range of R1 = R2 is 2 k ohm to 4 k ohm.

PHASE TRACKING THREE-PHASE GENERATOR



ELECTRONIC ENGINEERING

Fig. 92-8

Circuit Notes

Using a single chip LM324 can, with active R-C networks, reduce the size of a 3-phase waveform generator, and prove useful in compact and stable 3-phase inverters. One quarter of an LM324 is used as a Wien bridge oscillator generating a pure sinusoidal waveform while the remaining parts of the LM324 are used as three 120° fixed phase shifters. Initially potentiometer R3 should be varied to adjust the loop gain of the oscillator in order to start the oscillator.

LINE FREQUENCY SQUARE-WAVE GENERATOR

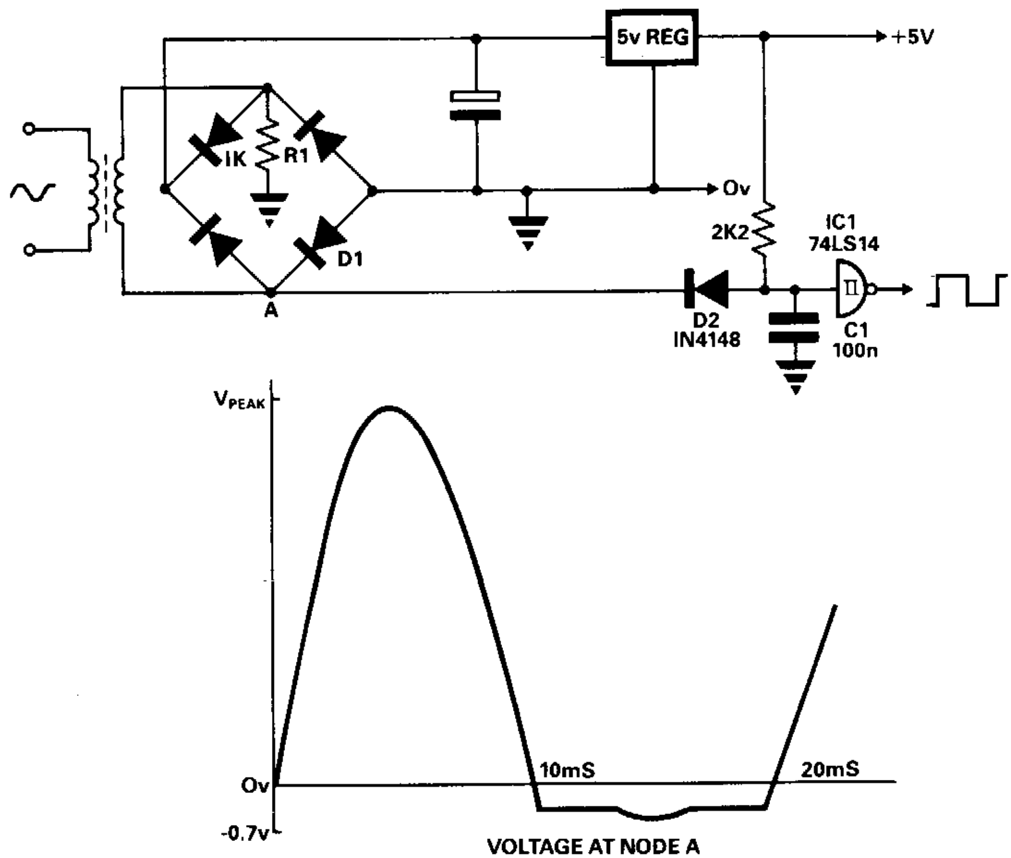


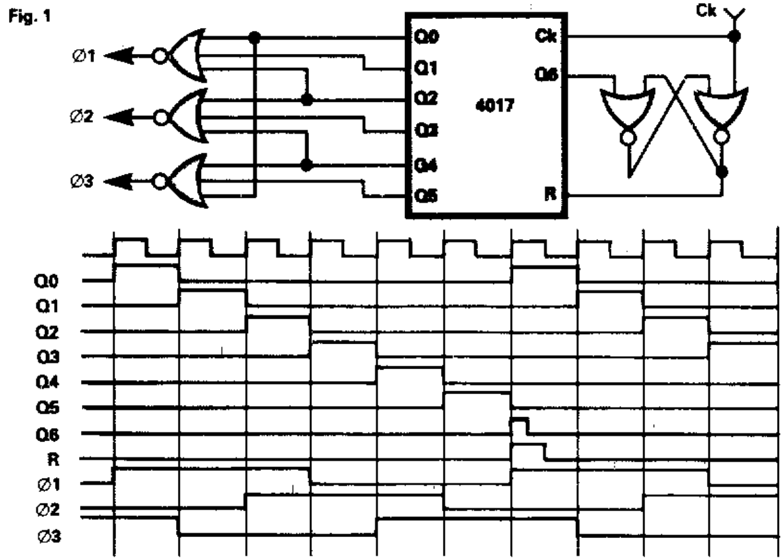
Fig. 92-9

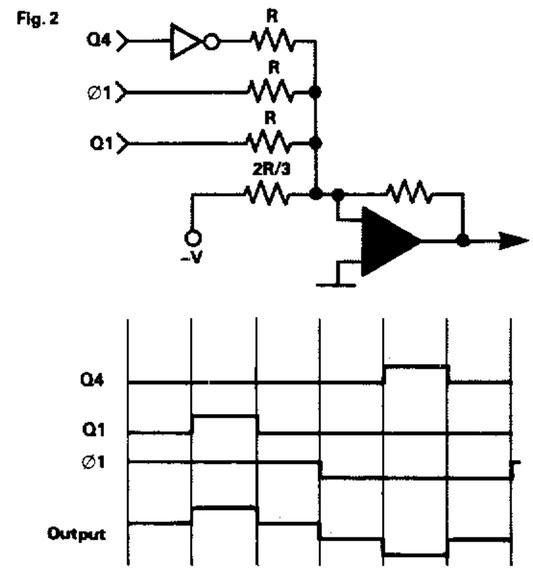
ELECTRONIC ENGINEERING

Circuit Notes

With only three components and a buffer, a line frequency square wave having a 1:1 duty cycle may be derived from the power supply. During the alternate half-cycle, however, A is effectively clamped to −0.7 V by D1 in the bridge which offsets the forward voltage across D2 giving an input to IC1 of approximately 0 V. When A rises above +5 V, D2 is reverse biased and remains at +5 V. R1 is needed to load the transformer secondary maintaining a distortion-free waveform at A during the time the diode bridge is not conducting. C1 although not essential may be required to remove transients.

THREE PHASE SQUARE-WAVE OUTPUT GENERATOR





ELECTRONIC ENGINEERING

Fig. 92-10

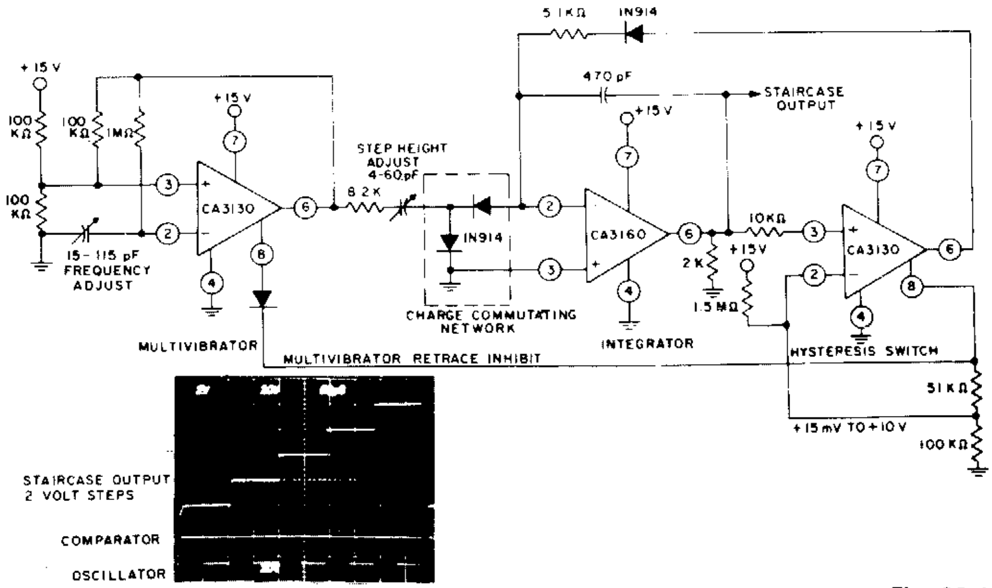
Circuit Notes

This circuit gives a 3 phase square-wave output for a variable speed motor drive. Operation is straightforward, the 4017 counter is synchronously reset after six clock inputs. The six outputs are combined to give the required waveforms. It is interesting to note that although NOR gates are shown, OR gates will give effectively the same result. The circuit can be extended as shown in Fig. 2 to give pseudo-sine waves if that is required, but that will diminish the simplicity of the circuit.

93
Staircase Generator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

STAIRCASE GENERATOR



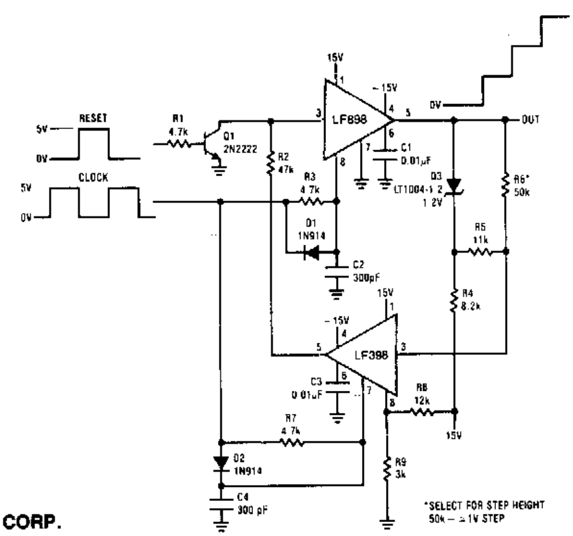
GENERAL ELECTRIC/RCA

Fig. 93-1

Circuit Notes

The circuit uses three BiMOS op amps. Two CA3130's are used, one as a multivibrator and the other as a hysteresis switch. The third amplifier, a CA3160, is used as a linear staircase generator.

STAIRCASE GENERATOR II



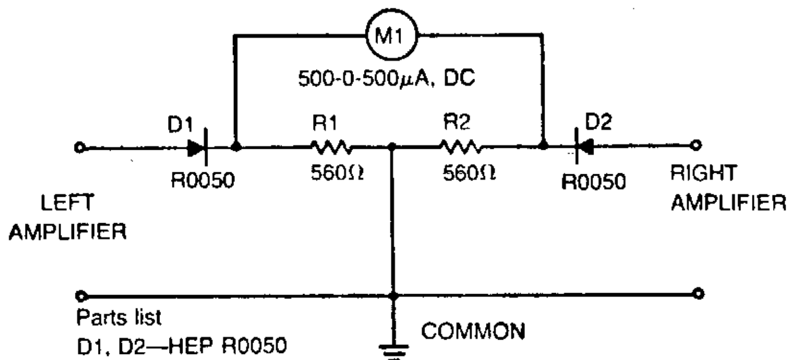
LINEAR TECHNOLOGY CORP.

Fig. 93-2

94
Stereo Balance Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

STEREO BALANCE TESTER



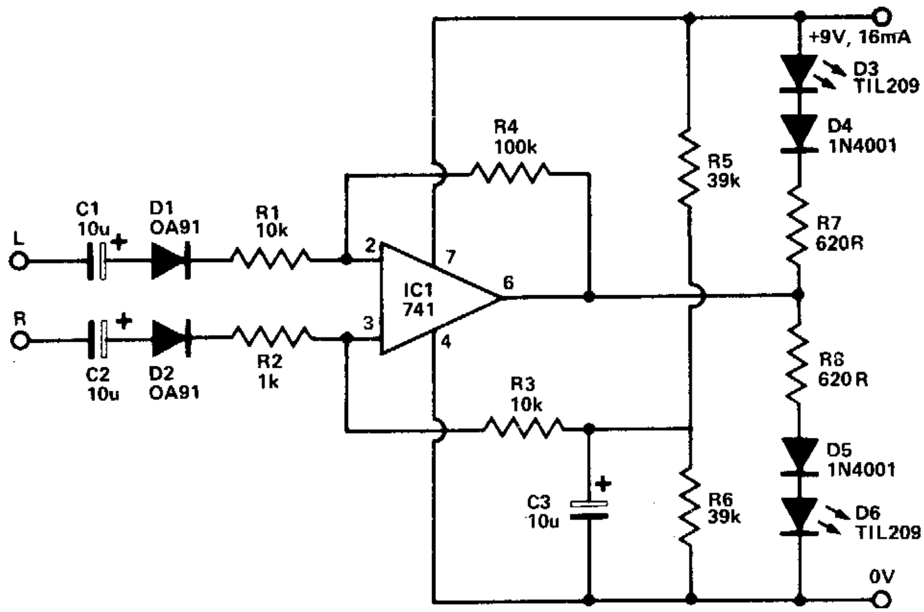
Parts list
D1, D2—HEP R0050
R1, R2—560Ω, 1/2W
M1—400-0-500μA dc meter, zero center scale Simpson panel,
Newark Electronics 55F480

TAB BOOKS INC. *Fig. 94-1*

Circuit Notes

The meter will show volume and tone control balance between left and right stereo amplifiers. For maximum convenience the meter is a zero-center type. Resistors are five percent or better and the diodes a matched pair. Optimum stereo level and phase balance occurs for matched speakers when the meter indicates zero. If the meter indicates either side of zero, the levels are not matched or the wires are incorrectly phased. Check phasing by making certain the meter leads are connected to the amplifier hot terminals and the common leads go to ground.

STEREO BALANCE METER



ELECTRONICS TODAY INTERNATIONAL

Fig. 94-2

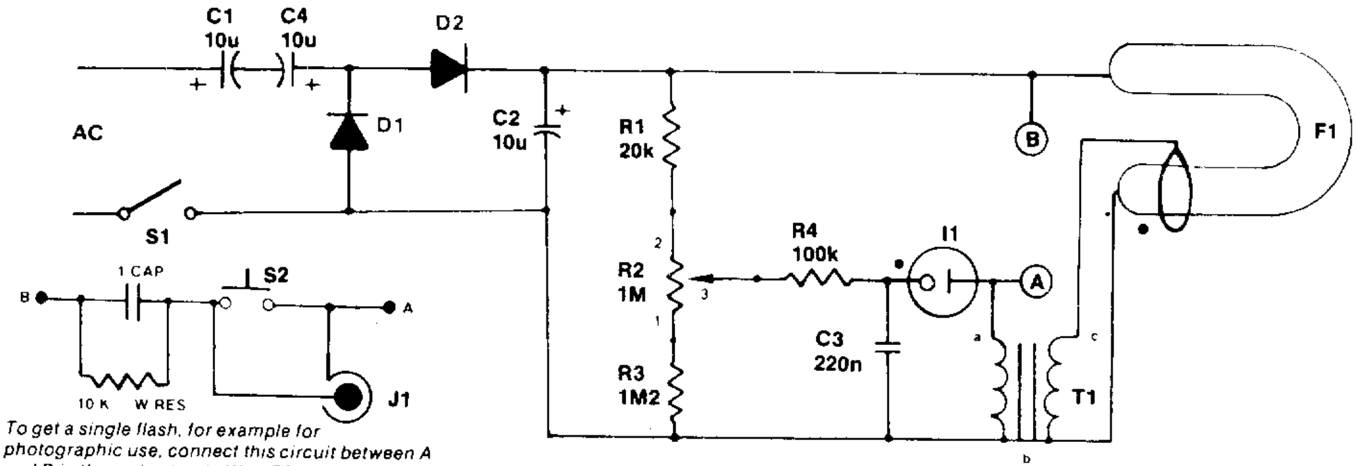
Circuit Notes

To use the indicator, switch the amplifier to mono mode and adjust the balance control until both LEDs are equally illuminated. The amplifier is now in perfect stereo mode balance.

95
Strobe Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SIMPLE STROBE



To get a single flash, for example for photographic use, connect this circuit between A and B in the main circuit. With R2 in the position for slowest flash rate (ie no flashes!) S2 will provide the desired single flash. Alternatively J1 can be used to allow the camera's flash contacts to trigger the strobe.

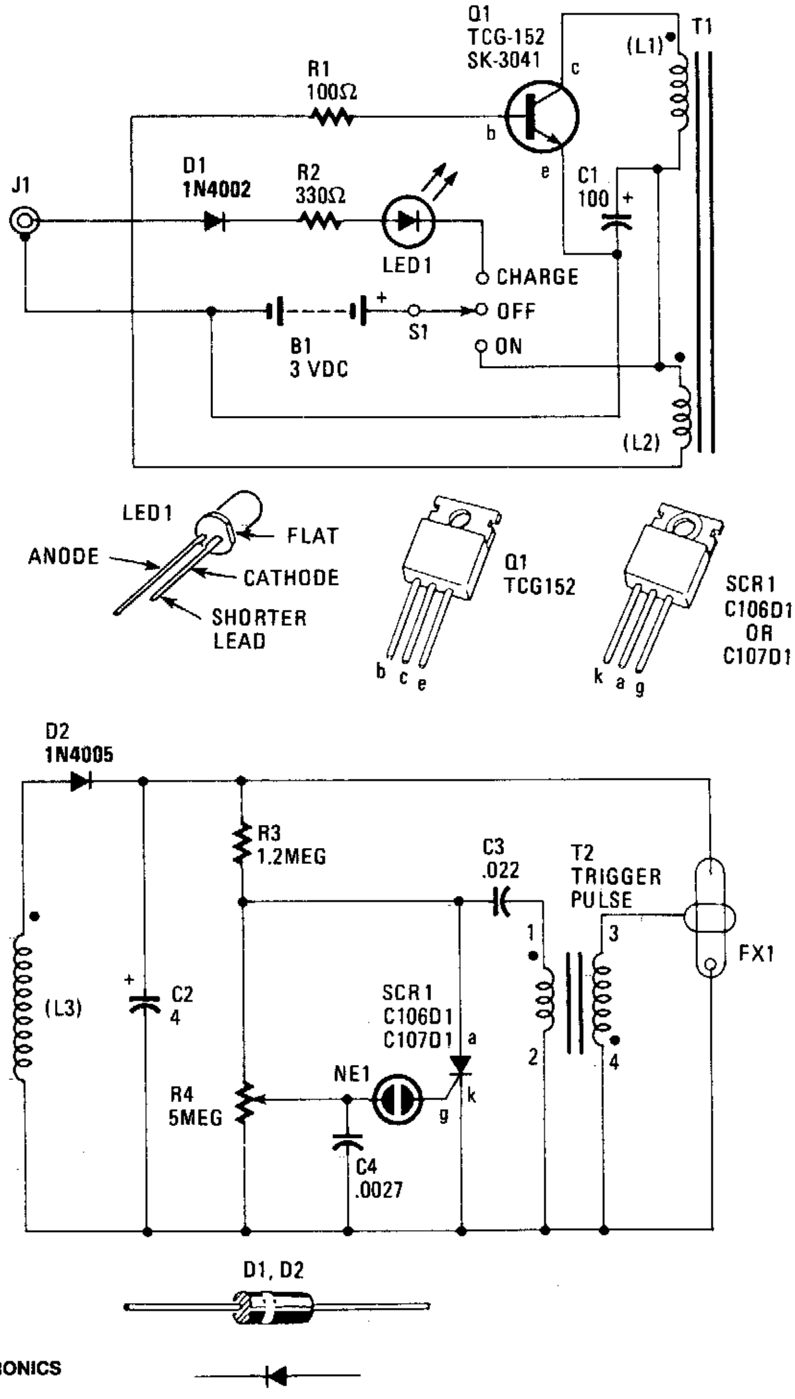
ELECTRONICS TODAY INTERNATIONAL

Fig. 95-1

Circuit Notes

Initially the neon and xenon lamps are not conducting and act like a very high (almost infinite) resistance. Capacitors C1 and C4 in conjunction with D1 and D2 form a voltage doubler circuit, which can charge C2 up to about 300 Vdc after several ac cycles. Voltage increases as current is supplied through R1 and R2. Neon bulb I1 will all of a sudden start to conduct when the voltage across C3 reaches I1's ionization potential. While conducting, the resistance of the bulb will be relative low. Due to this sudden conduction, a pulse of current will pass through the primary of T1. The turns ratio is such that about 400 V will be developed at the secondary. The xenon tube is similar to the neon bulb in that it produces light when the gas ionizes and conducts. However, it is designed so that an external signal (the 4 kV on the metal ring around the tube) ionizes the gas and initiates the conduction. When F1 conducts, it discharges C2. At this point, the whole cycle starts over again. The purpose of R2 is to vary the rate at which C3 charges, and hence the repetition rate of the strobe.

SAFETY FLARE



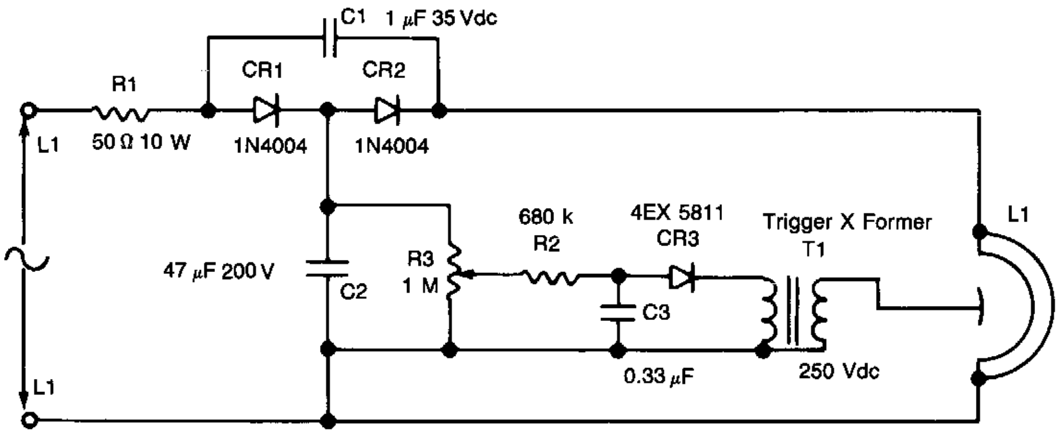
HANDS-ON ELECTRONICS

Fig. 95-2

Circuit Notes

When S1 is on, power is applied to an oscillator composed of Q1, R1, C1, L1, and L2. Coil L1 is the primary winding of T1, and L2 is the feedback winding. When Q1 turns on, its collector current saturates T1's ferrite core. That, in turn, removes the base drive to Q1 through L2. Transistor Q1 then turns off. As the field around L1 and L2 decays, Q1 will eventually turn on again, and the cycle repeats over, and over. Transformer T1 is a step-up, ferrite-core, potted-type unit whose secondary-winding (L3) output is rectified by D2 and filtered by C2. That capacitor charges up to around 250 to 300 volts, which is applied to the resistor divider composed of R3 and R4, along with the flash tube FX1. Capacitors C3 and C4 will charge up to around 200 and 100 volts, through R3 and R4, respectively. Flash rate is adjustable via R4. When the charge on C4 gets to around 100 volts, neon lamp NE1 fires discharging C4 into the gate circuit of silicon control rectifier SCR1. The SCR1 turns on discharging C3 into the primary winding of trigger-pulse transformer T2. Transformer T2 is another step-up, pulse-type unit providing an output of around 4 kW across transformer T2's secondary winding. The xenon gas inside FX1 is ionized and a bright flash is emitted. Finally, C3 quickly discharges through L4, and the cycle repeats over, and over.

DISCO-STROBE LIGHT



WILLIAM SHEETS

Fig. 95-3

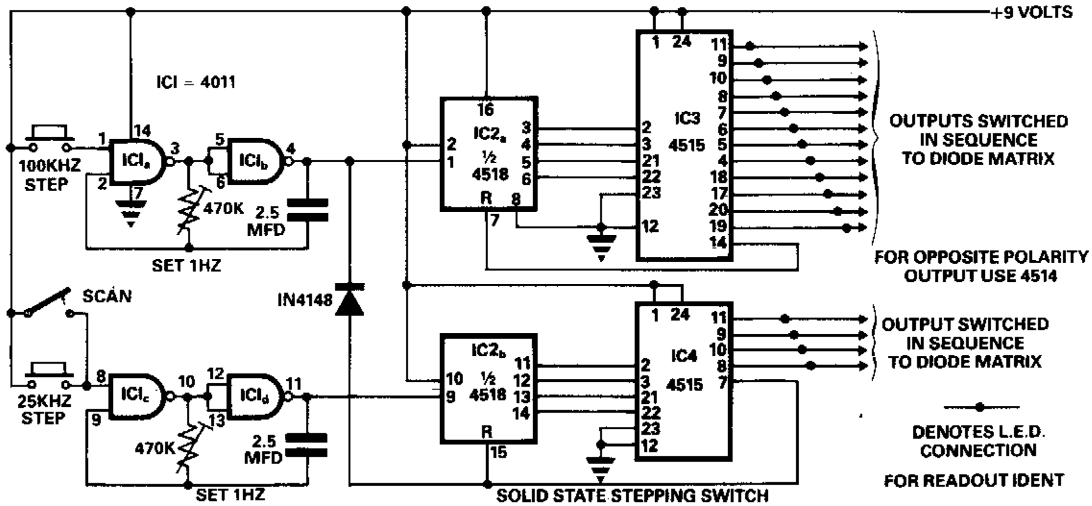
Circuit Notes

This circuit uses a voltage doubler CR1 and CR2 to obtain about 280 V dc across C1. C2 and R3 form a voltage divider to obtain a dc voltage to change C3 thru R2. When CR3 fires, a high voltage is generated in T1, firing L1.

96
Switch Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SOLID STATE STEPPING SWITCH



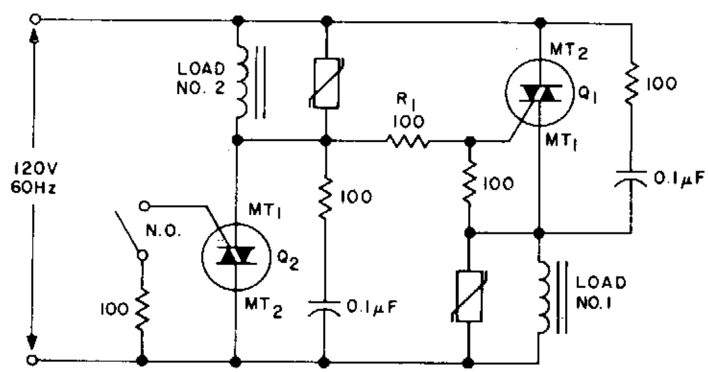
ELECTRONIC ENGINEERING

Fig. 96-1

Circuit Notes

This circuit was designed to make switching of a 48-channel mobile transceiver safe to operate while mobile. The oscillators allow for single-stepping or a scanning function. The scan facility allows for stepping through all 48 channels to check for occupancy or otherwise, and each output is indicated with an LED and labeled accordingly, so at-a-glance indication is possible. With full scope of this circuit it is possible to scan 256 channels and by adding more 4 to 16 line encoders etc. you could switch to any required number.

AC-STATIC SPDT SWITCH



GENERAL ELECTRIC

Fig. 96-2

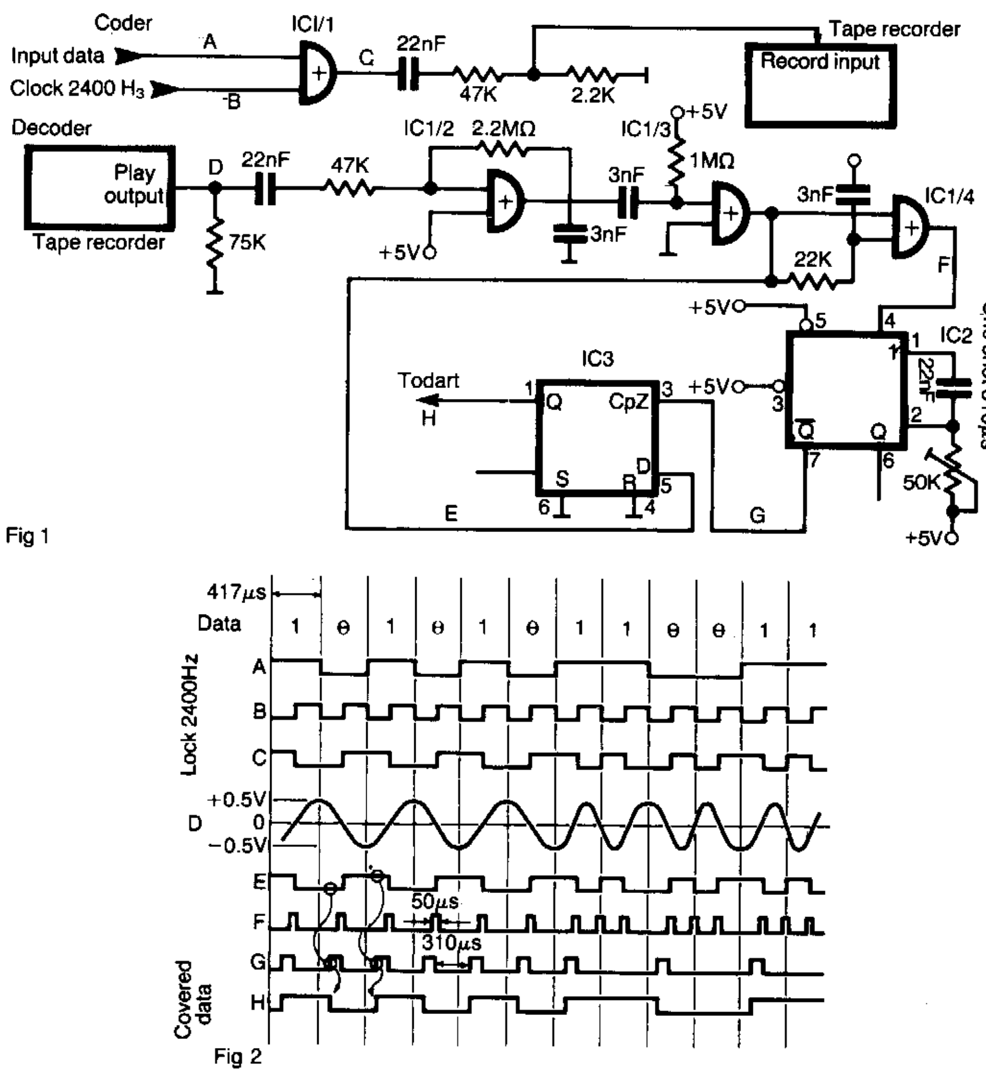
Circuit Notes

An SPDT solid state relay is shown. When voltage is applied Q1 will turn on, activating load #1, because the full line voltage appears across Q2, supplying gate current through R1. When S1 is closed, Q2 turns on removing the gate drive from Q1 and activating load #2.

97
Tape Recorder Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TAPE RECORDER INTERFACE



ELECTRONIC ENGINEERING

Fig. 97-1

Circuit Notes

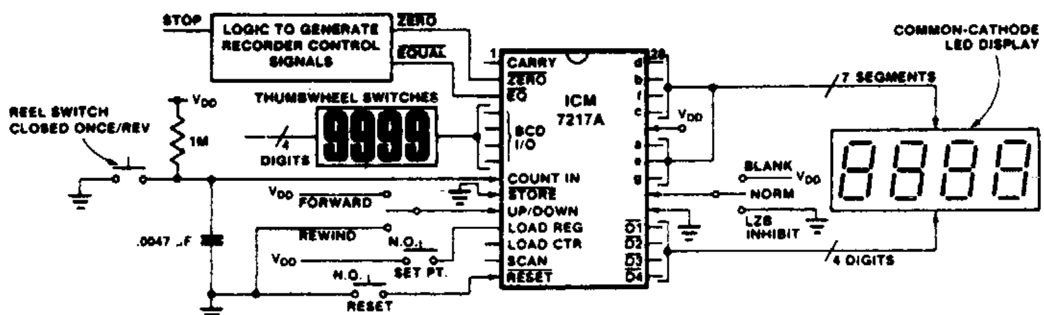
The interface allows data to be saved on an ordinary tape recorder at a speed of 2400 bit/s.

The serial stream of data Fig. 1 (A) is coded with a clock of 2400 Hz (B), by means of XOR gate IC 1/1. Logical "high" and "low" appear as shown in Fig. 2 (C). These impulses are lowered in amplitude and feed into the record input of a low cost tape recorder.

TAPE RECORDER INTERFACE, Continued.

During the playback, pulses (D) are amplified with CMOS gate IC 1/2 connected as a linear amplifier, and providing a TTL level signal shown in (E). On both positive and negative transitions IC 1/4 forms short pulses as shown in (F) (approx. 50 μs) that triggers one shot IC2. A monostable one shot pulse width is adjusted to be ¾ of bit length (310 μs). A change from "high" to "low" in a coded stream generates a "low" pulse width of one bit cell. The same is for change from "low" to "high" that generates a "high" pulse of the same width. During this pulse one shot latches the state of line E in D type flip-flop IC3 (G). When a stream consists of multiple "ones" or "zeros," the one shot is retriggered before it comes to the end of the quasistable state and the state of the flip-flop remains unchanged. The original data stream is available at the output of the flip-flop (H). Z80 the DUART that receives these pulses is programmed so that the receiver clock is 16 times the data rate (38.4 kHz).

TAPE RECORDER POSITION INDICATOR/CONTROLLER



INTERSIL

Fig. 97-2

Circuit Notes

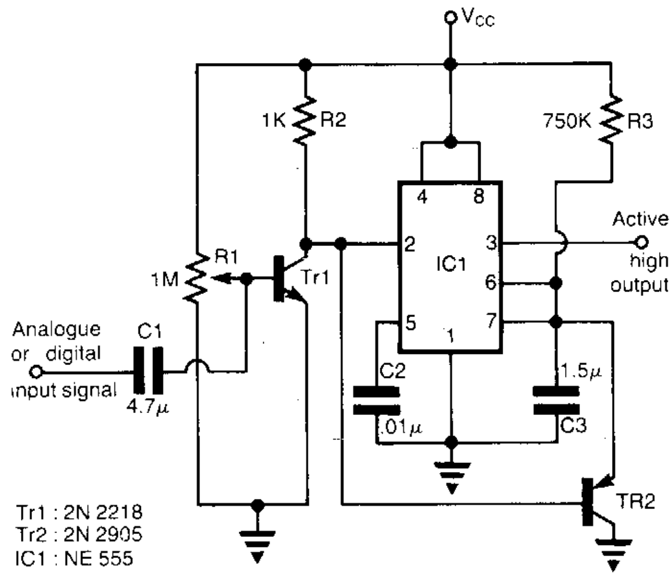
This circuit is representative of the many applications of up/down counting in monitoring dimensional position. In the tape recorder application, the LOAD REGISTER, EQUAL, and ZERO outputs are used to control the recorder. To make the recorder stop at a particular point on the tape, the register can be set with the stop at a particular point on the tape, the register can be set with the stop point and the EQUAL output used to stop the recorder either on fast forward, play or rewind.

To make the recorder stop before the tape comes free of the reel on rewind, a leader should be used. Resetting the counter at the starting point of the tape, a few feet from the end of the leader, allows the ZERO output to be used to stop the recorder on rewind, leaving the leader on the reel. The 1 M ohm resistor and .0047 μF capacitor on the COUNT INPUT provide a time constant of about 5 ms to debounce the reel switch. The Schmitt trigger on the COUNT INPUT of the ICM7217 squares up the signal before applying it to the counter. This technique may be used to debounce switchclosure inputs in other applications.

98
Telephone-Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

SPEECH ACTIVITY DETECTOR FOR TELEPHONE LINES



ELECTRONIC ENGINEERING *Fig. 98-1*

Circuit Notes

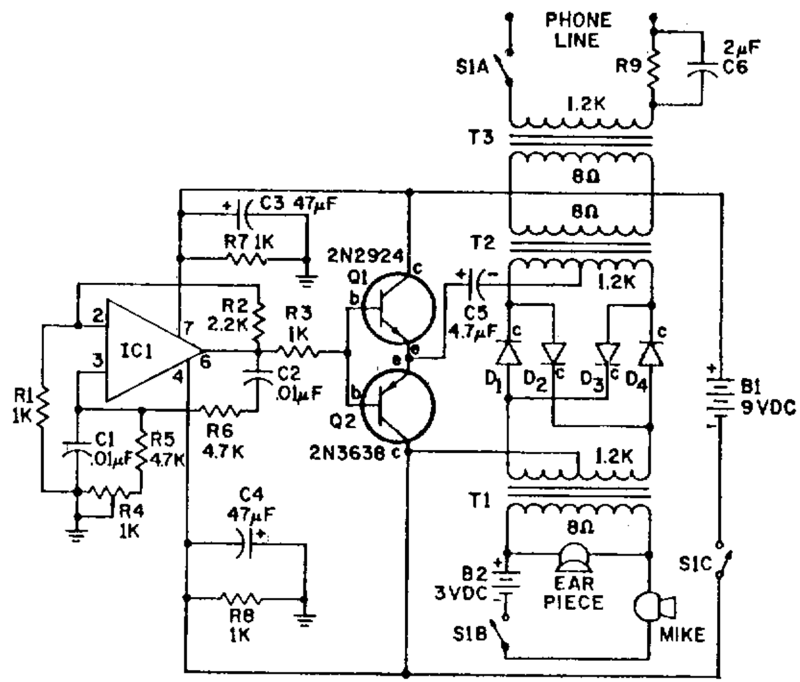
The circuit can be used in telephone lines for speech activity detection purposes. This detection is very useful in the case of half-duplex conversation between two stations, in the case of simultaneous transmission of voice and data over the same pair of cables by the method of interspersion data on voice traffic, and also in echo suppressor devices. The circuit consists of a class-A amplifier in order to amplify the weak analog signals (in the range 25-400 mW of an analog telephone line).

The IC1 is connected as a retriggerable monostable multivibrator with the Tr2 discharging the timing capacitor C3, if the pulse train reaches the trigger input 2 of IC1 with period less than the time:

$$T_{high} = 1.1\ (R3\ C3)$$

The output 3 of IC1 is active ON when an analog or digital signal is presented at the output and it drops to low level, T_{high} seconds after the input signal has ceased to exist.

SCRAMBLE PHONE



B1—9-volt battery, Eveready 216 or equiv.
B2—3-volt battery, two AA penlight cells in series
C1, C2—0.01 uF polystyrene capacitor, 100 VDC or better
C3, C4—47 uF electrolytic capacitor, 25 VDC or better
C5—4.7 uF electrolytic capacitor, 25 VDC or better
C6—2 uF paper or mylar capacitor, 50 VDC or better
D1 to D4—Diode, IN914, HEP-156
IC1—Integrated circuit, Signetics N5741K or equiv.
Q1—NPN transistor, 2N2924, HEP-724
Q2—PNP transistor, 2N3638, HEP-716
R1, R3, R7, R8—1000-ohm, ½-watt resistor
R2—2,200-ohm ½-watt resistor
R4—1000-ohm potentiometer
R5, R6—4,700-ohm, ½-watt resistor
R9—Limit line current to 25mA (see text)
S1A, S1B, S1C—Phone hook switch (see text)
T1 to T3—Small transistor audio transformer; 8-ohm primary, 1,200-ohm center taped secondary.
Misc.—Surplus telephone (see Lafayette, Radio Shack, EDI, BA catalogs), battery holders, hardware, knob, wire, solder, etc.

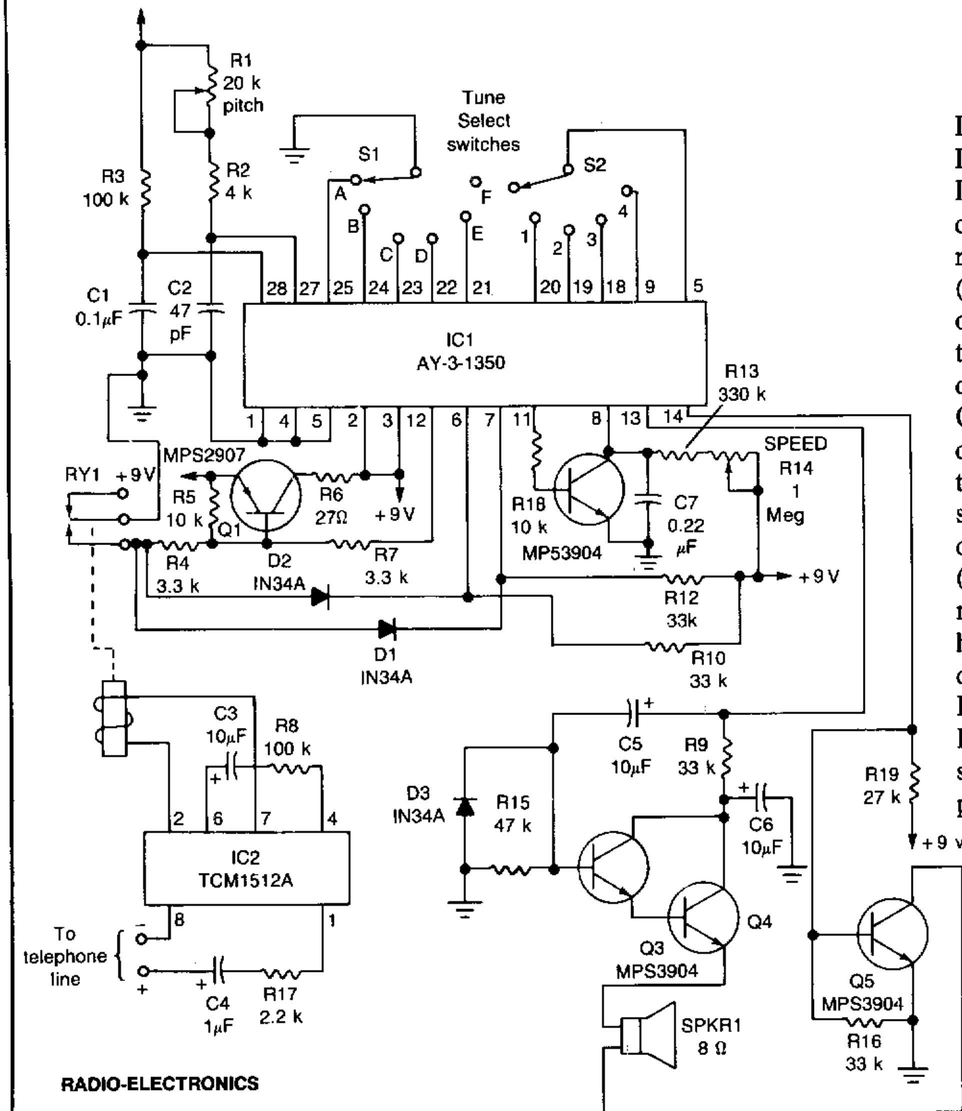
TAB BOOKS, INC.

Fig. 98-2

Circuit Notes

IC-1 and the associated circuitry form a stable audio tone generator that feeds a buffer amplifier, Q1 and Q2. The tone output is taken from the emitters of the transistor pair to supply a carrier voltage for a balanced modulator made up of four diodes—D1 through D4—and T1 and T2. If the two transformers and the four diodes are perfectly matched (which is almost impossible to achieve and not necessary in any case) no carrier will appear at the input or output of T1 or T2. In a practical circuit, a small amount of unbalance will occur and produce a low-level carrier tone at the input and output of the balanced modulator. A telephone carbon mike and earpiece are connected to the low impedance winding of T1, with a three volt battery supplying the necessary mike current. Trim potentiometer R4 is used to make a fine frequency adjustment of the oscillator so that two scrambler units may be synchronized to the same carrier frequency. Rg limits line current to 25 mA.

MUSICAL TELEPHONE RINGER



Circuit Notes

The heart of the circuit is IC1, General Instrument's AY-3-1350 melody-synthesizer IC. IC2 is a TCM1512 telephone ring detector IC that is powered by the telephone line. The circuit's operation begins when IC2 senses a ring pulse on the telephone line. The detector (internally) rectifies the ring signal and then outputs a voltage to relay RY1 (an SPST reed-type relay with 5 volt contacts), causing its contacts to close. That pulls pin 12 (the ON/OFF control) of IC1 low (logic "0"), causing it to output a signal—the selected tune—to transistor amplifier Q2. The amplified signal is then fed to the speaker. The melody continues to play either until the tune is finished (at which time IC1 returns to the standby mode), or until someone takes the phone off the hook. Taking the phone off the hook discontinues the ring pulses to IC2, which opens RY1. When the relay contacts open, pin 12 of IC1 goes high, returning the circuit to the standby mode to wait for the next incoming phone call.

Fig. 98-3

DUAL TONE DECODING

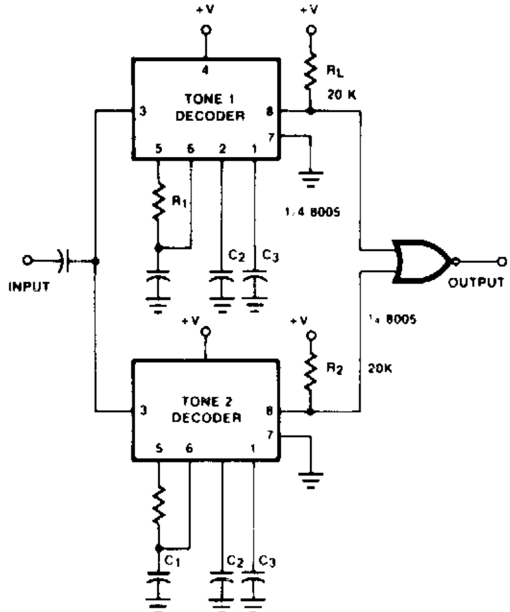


Figure 1A. Detection of Two Simultaneous or Sequential Tones

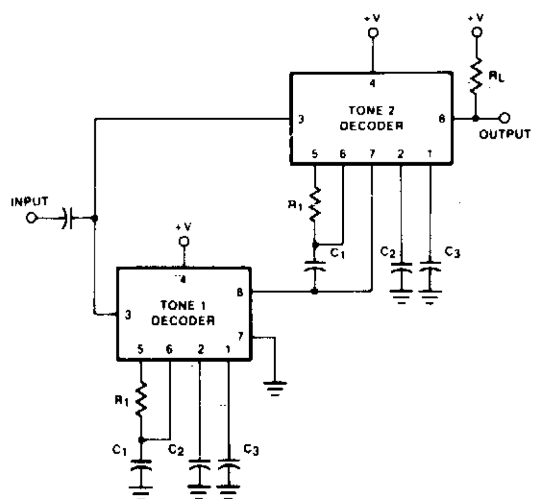


Figure 1C

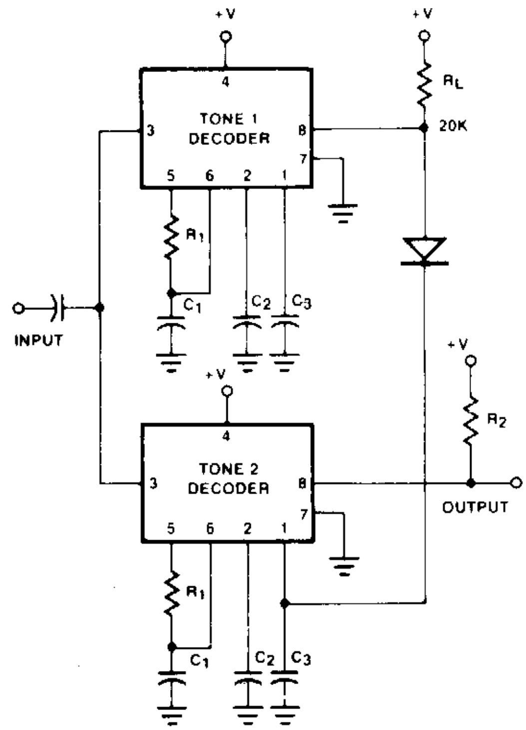


Figure 1B

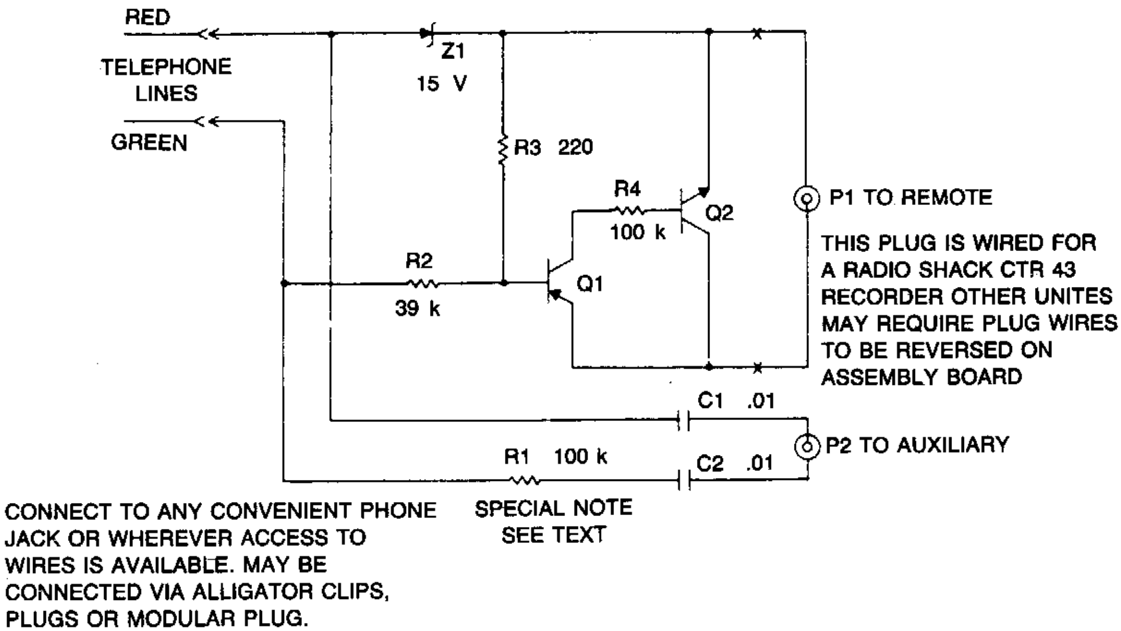
EXAR

Fig. 98-4

Circuit Notes

Two integrated tone decoders, XR-567 units, can be connected (as shown in Fig. 1A) to permit decoding of simultaneous or sequential tones. Both units must be on before an output is given. R1C1 and R′1C′1 are chosen, respectively, for Tones 1 and 2. If sequential tones (1 followed by 2) are to be decoded, then C3 is made very large to delay turn-off of Unit 1 until Unit 2 has turned on and the NOR gate is activated. Note that the wrong sequence (2 followed by 1) will not provide an output since Unit 2 will turn off before Unit 1 comes on. Figure 1B shows a circuit variation which eliminates the NOR gate. The output is taken from Unit 2, but the Unit 2 output stage is biased off by R2 and CR1 until activated by Tone 1. A further variation is given in Fig. 1C. Here, Unit 2 is turned on by the Unit 1 output when Tone 1 appears, reducing the standby power to half. Thus, when Unit 2 is on, Tone 1 is or was present. If Tone 2 is now present, Unit 2 comes on also and an output is given. Since a transient output pulse may appear at Unit 1 turn-on, even if Tone 2 is not present, the load must be slow in response to avoid a false output due to Tone 1 alone. The XR-267 Dual Tone Decoder can replace two integrated tone decoders in this application.

AUTOMATIC TELEPHONE RECORDING DEVICE

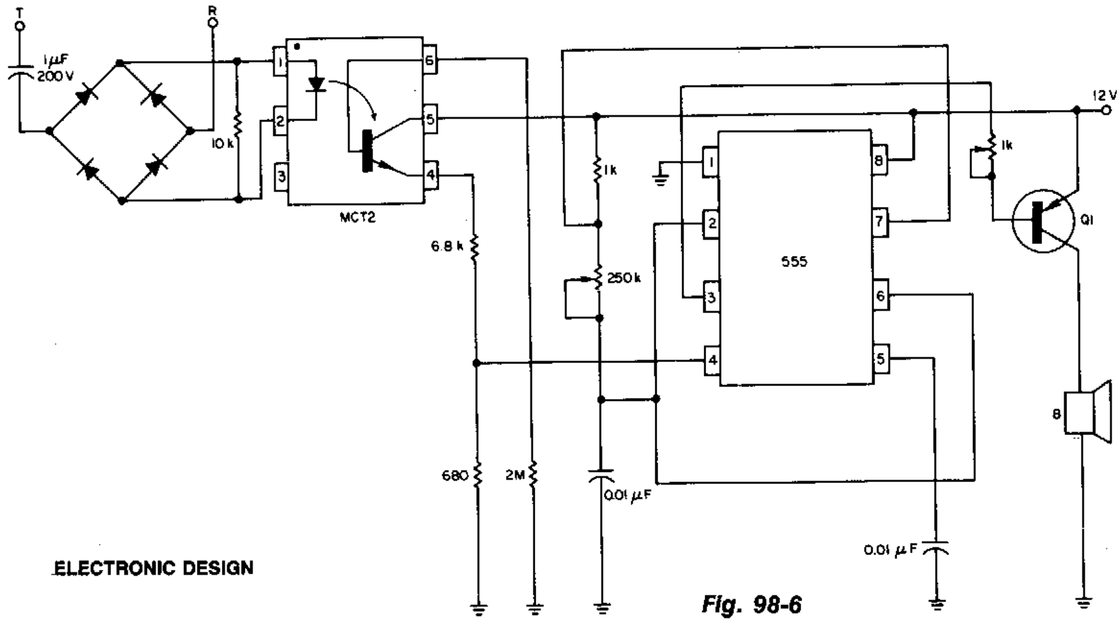


TAB BOOKS, INC. *Fig. 98-5*

Circuit Notes

The device is a dc switch that is normally on via the forward biasing of Q1 via R3. Q1 now clamps Q2 into a forward state by biasing its complement well into a saturated state via R4. The dc switch is turned off via a negative voltage above that of the zener (D1). This voltage is usually about 48 and is the on-hook value of the phone line. This negative voltage overrides the effect of R3 and keeps the circuit "off." When the phone is off the hook, the 48 volts drops to 10 volts, that is below the zener voltage of D1 and R3 now turns the circuit on. The audio signal is via attenuator resistor R1 and dc isolating capacitors C1, C2. The device is a high impedance switch that isolates the recording controlled device from the phone line via some relatively simple electronic circuitry. It requires no battery and obtains power for operating via the remote jack that in most recorders is a source of 6 volts. When clamped to ground it initiates recorder operation. The unit interfaces with most portable cassette recorders providing they contain a remote control jack.

TELEPHONE RINGING DETECTOR, FREQUENCY AND VOLUME CONTROLLED



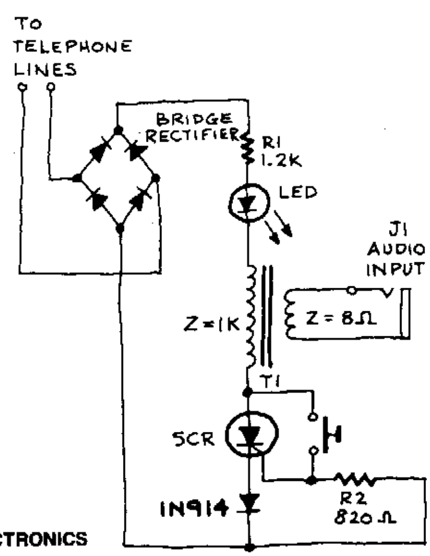
ELECTRONIC DESIGN

Fig. 98-6

Circuit Notes

With the 555 timer connected as a multivibrator and an opto-isolator, a remote speaker can be driven.

MUSIC ON HOLD



RADIO-ELECTRONICS

Circuit Notes

With this music-on-hold device, you can answer the phone in one room, place the caller on hold, and then pick up the phone again at another location. When you pick up the phone the second time, you automatically deactivate the music-on-hold feature and can continue your conversation.

Fig. 98-7

CIRCUIT MONITORS BLINKING PHONE LIGHTS

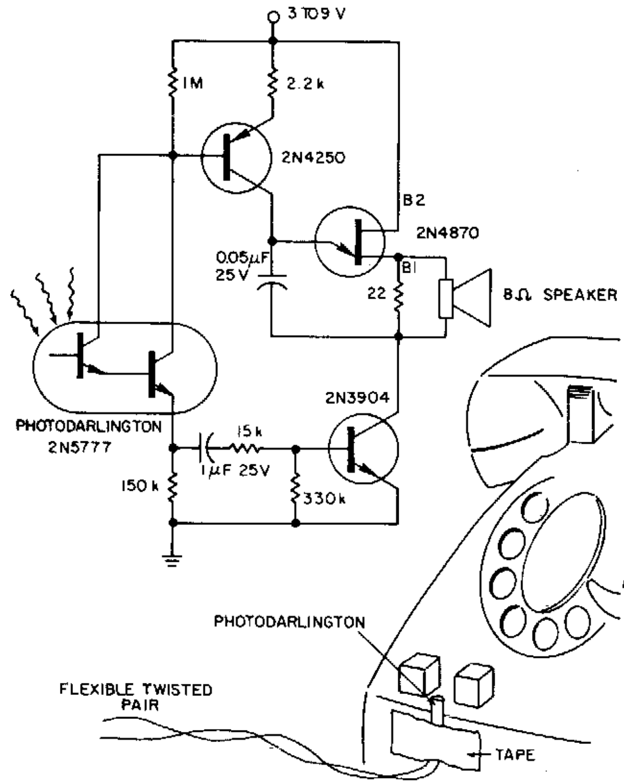


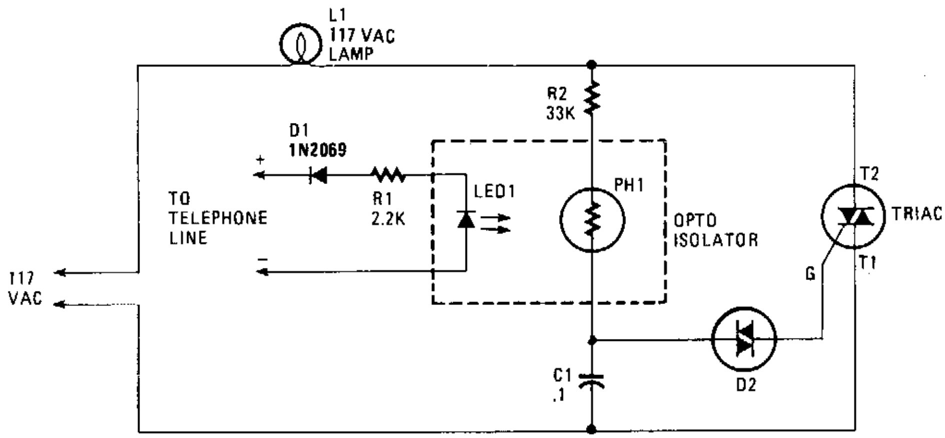
Fig. 98-8

ELECTRONIC DESIGN

Circuit Notes

A 2N5777 photo-Darlington cell picks up blinking light from the transparent plastic buttons. The power is switched ON and OFF by a hi-beta 2N3904 transistor. The circuit's 9 V battery can be left continuously connected. Less than a micro-ampere is drawn—even with normal, office ambient light and the phone lights not flashing. For noisy locations, the tone can be made louder with an output transformer (ratio of 250:8) or a 100 ohm speaker that replaces the 22 ohm resistor in the output.

PHONE LIGHT

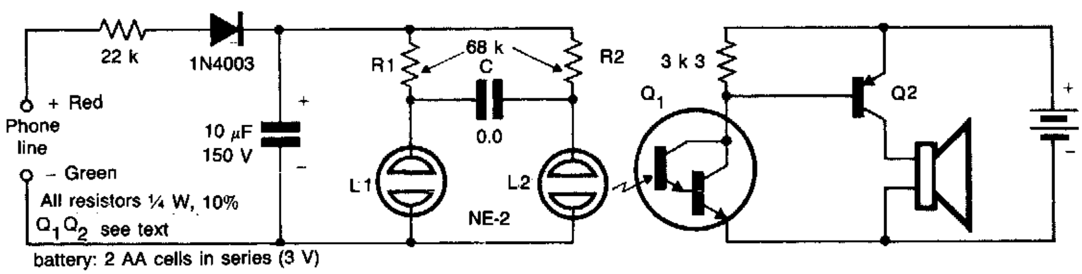


HANDS-ON ELECTRONICS *Fig. 98-9*

Circuit Notes

When the phone does ring the triac is triggered into conduction by a signal applied to its gate (G) through a bilateral switch (diac), D2. The triac acts as a switch, conducting only when a signal is present at the gate.

HIGH ISOLATION TELEPHONE RINGER

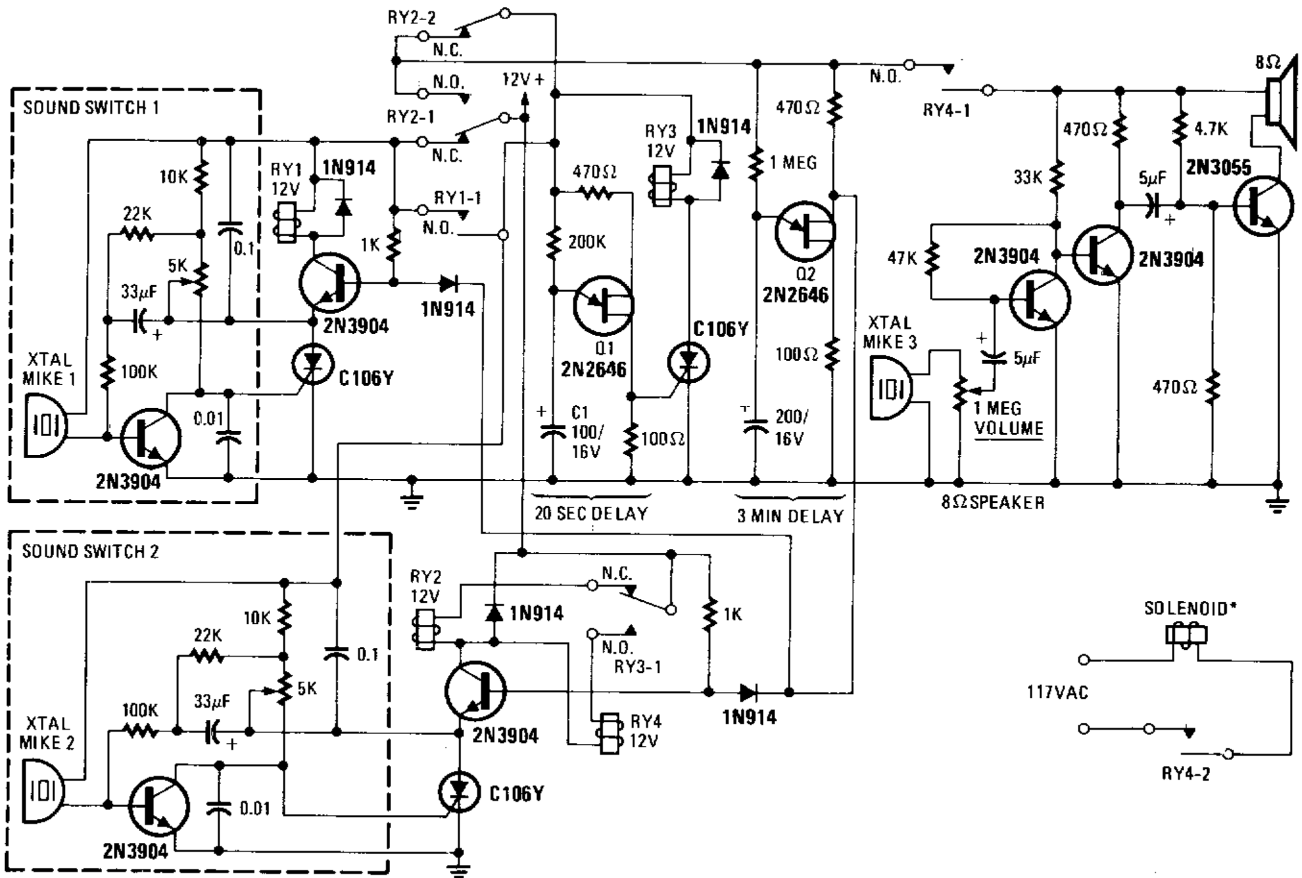


ELECTRONIC ENGINEERING *Fig. 98-10*

Circuit Notes

The diode rectifies the ringing signal to supply the operating power to the audio relaxation oscillator made up of L1, L2, R1, R2, and C. Moreover, L2 together with Q1 acts as an opto-isolator, totally isolating the telephone line from the rest of the circuit. The oscillator audio frequency is optically coupled to the photo-Darlington which drives Q2 and thus the speaker. The 10 μF capacitor is not large enough to smooth the ringing ripple completely. This results in frequency modulation of the audio oscillator giving it an attention-getting warble.

REMOTE TELEPHONE MONITOR



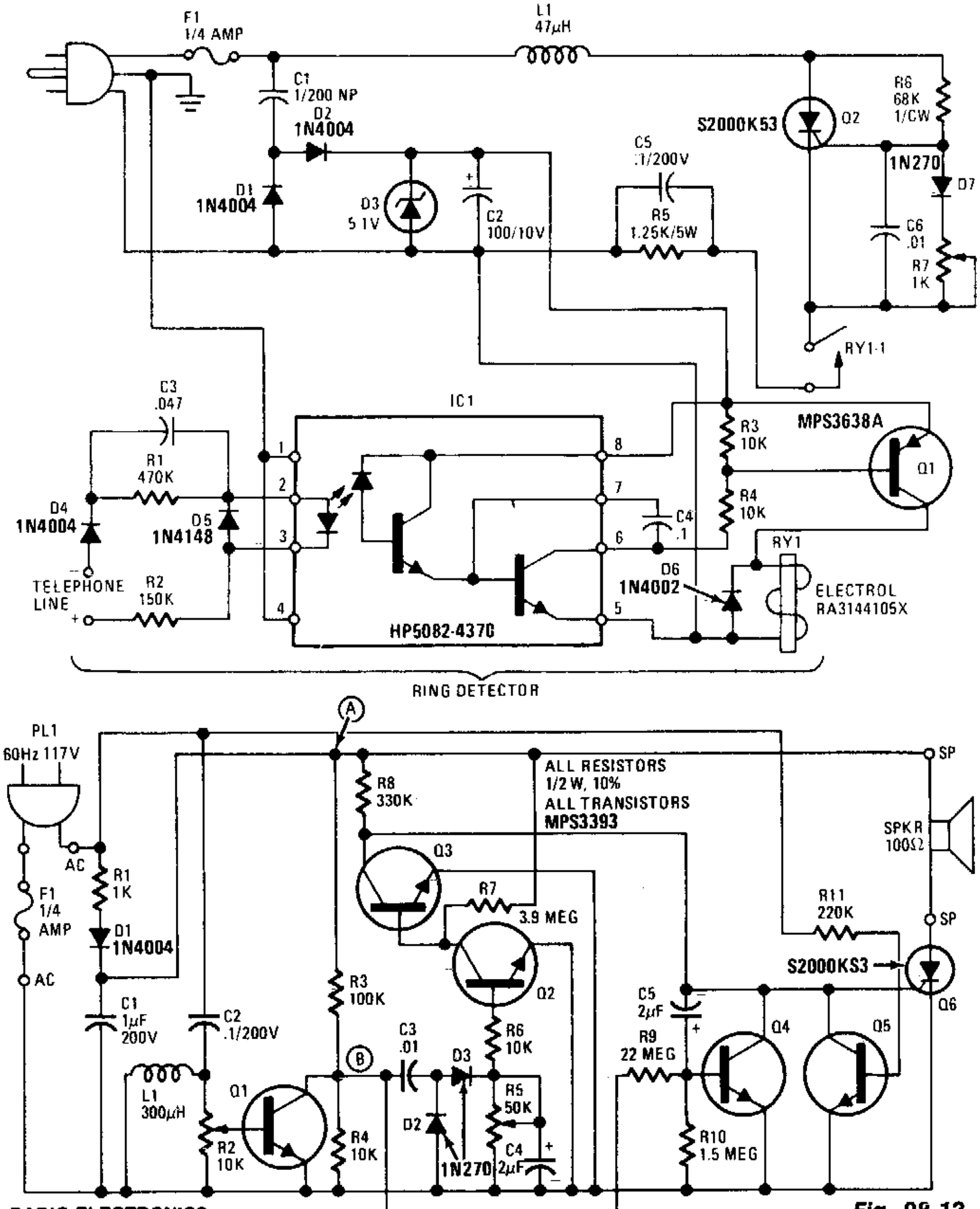
RADIO-ELECTRONICS

Fig. 98-11

Circuit Notes

This device monitors sounds in home or office when a telephone is called from a remote location.

PLUG-IN REMOTE TELEPHONE RINGER



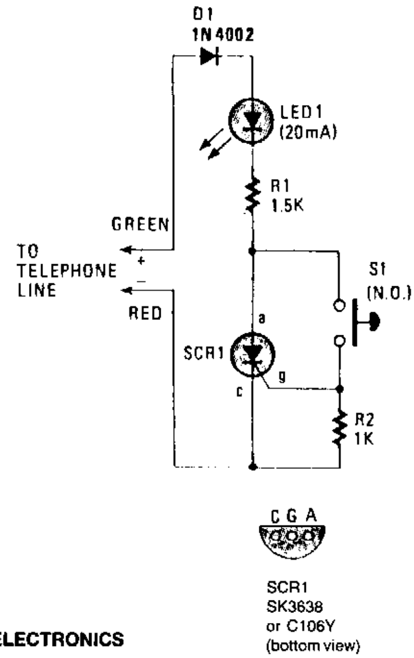
RADIO-ELECTRONICS

Fig. 98-12

Circuit Notes

This device consists of a ring detector connected to the telephone line. When the telephone rings, the ring detector impresses high-frequency pulses on the ac power line. A receiver placed anywhere on the same power line detects these pulses and emits an audible tone in synchronization with the telephone signal.

TELEPHONE HOLD BUTTON

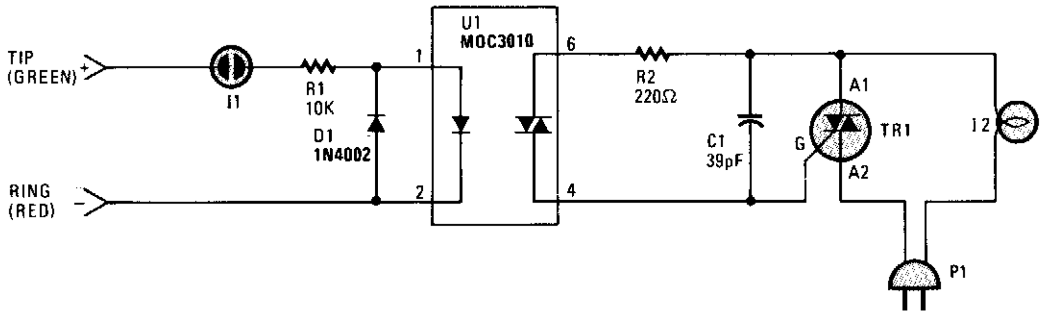


HANDS-ON ELECTRONICS *Fig. 98-13*

Circuit Notes

The on-hook (no load) voltage across the red-green wires will be 48 V or slightly less when all telephones are on-hook (disconnected). When any telephone goes off-hook the load current flowing in the telephone causes the voltage to fall below 5 volts dc. Although the telephone hold is connected across the red-green wires, silicon control rectifier SCR1 is open; so there is no current path across the telephone line. To hold the call, depress normally-open switch S1 and hang up the telephone (still depressing S1). When the phone goes on-hook the red-green voltage jumps to 48 volts dc. Since switch S1 is closed, a positive voltage is applied to SCR1's gate, which causes SCR1 to conduct, thereby completing the circuit across the telephone line through D1, LED1, R1, and SCR1. The current that flows through those components also causes the LED to light up—indicating that the telephone line is being held. The effective load across the red-green wires is the 1500 ohm value of R1, which is sufficient to seize the line while limiting the current through the LED to a safe value. When the telephone, or an extension, is once again placed off-hook the red-green voltage falls to 5 volts or less. But diode D1 has a normal voltage drop—called the breakover voltage—of 0.7 volts, and the LED has a forward drop of 2.0 volts. Excluding the voltage drop across R1 there is a maximum of 2.3 volts available for SCR1, which is too low to maintain conduction; so SCR1 automatically opens the hold circuit when any telephone goes off-hook.

TELEPHONE BLINKER

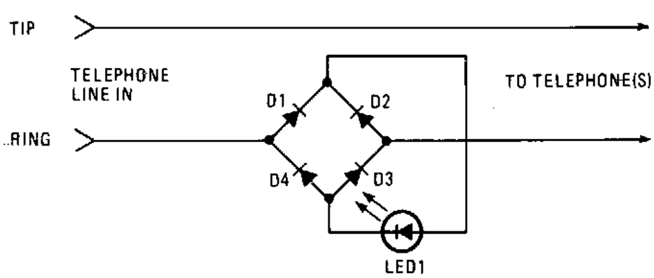


HANDS-ON ELECTRONICS *Fig. 98-14*

Circuit Notes

A small neon lamp is triggered into conduction by the telephone's ringing voltage, passes just enough current to activate the LED in optocoupler U1, which in turn triggers the 6-A Triac that controls I2—a 117-Vac lamp or bell. (Capacitor C1 is necessary only when the circuit is used to drive a bell.) The lamp will flash off-and-on at the ringing rate, which is normally around 20 Hz. If a 117 Vac bell is used, connect it in place of the lamp.

TELEPHONE "IN USE" INDICATOR

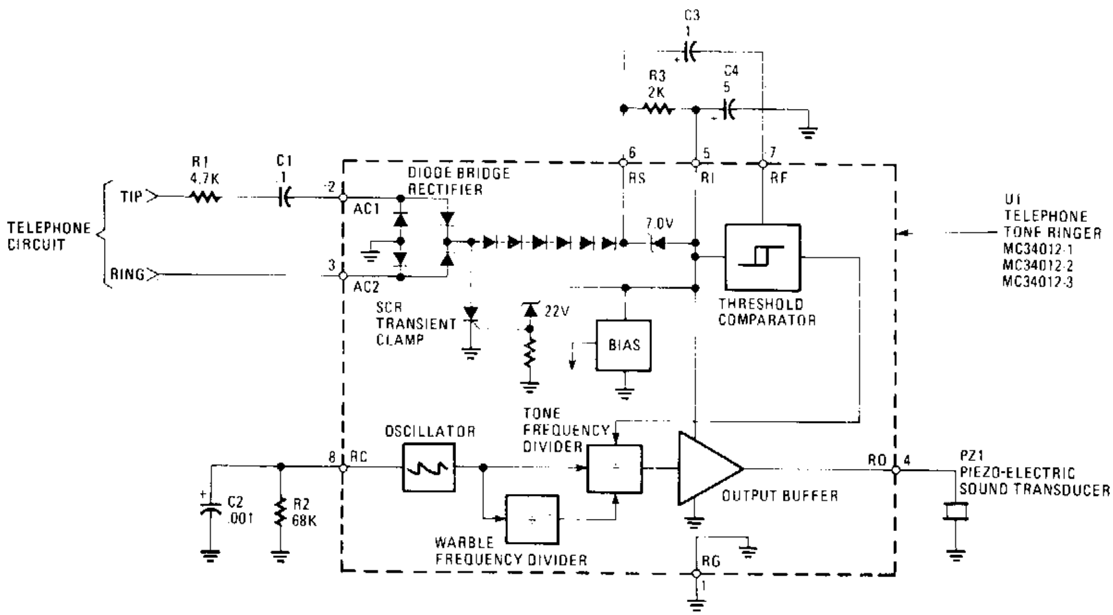


HANDS-ON ELECTRONICS *Fig. 98-15*

Circuit Notes

This circuit functions as a line-current sensor and can be connected in series with either of the phone lines. For the circuit to indicate an "in use" status for all phones on a single line, it must be connected in series with the phone line before, or ahead of all phones on the line. Since the power for the circuit is supplied by the phone company, a circuit could be added to each phone as an off-hook indicator.

TONE RINGER

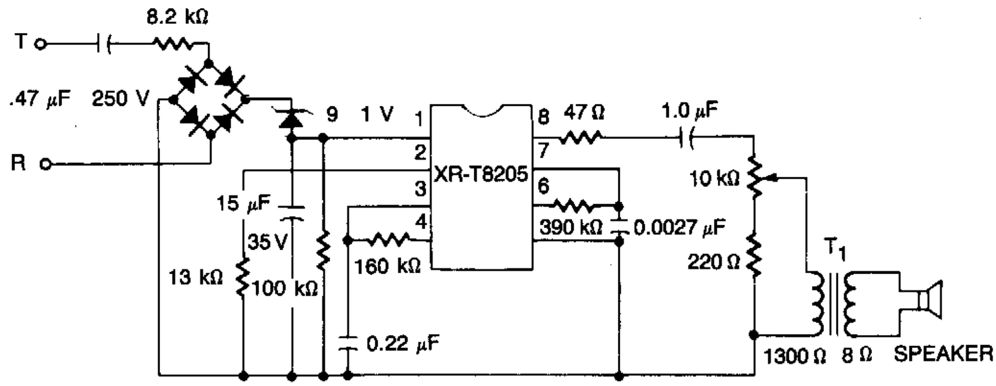


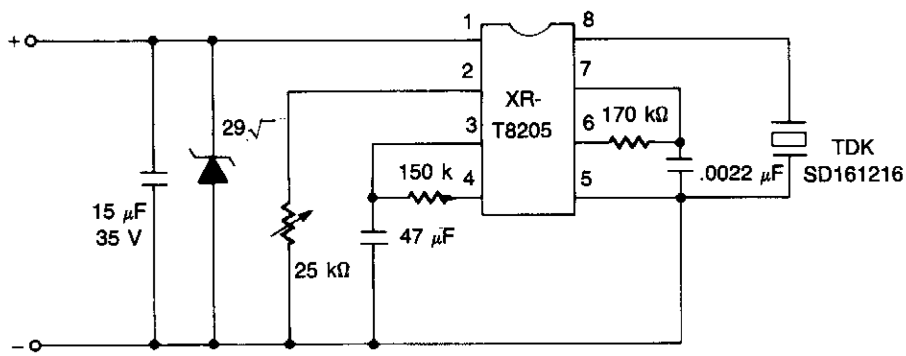
HANDS-ON ELECTRONICS *Fig. 98-16*

Circuit Notes

The MC34012 tone-ringer chip derives its power by rectifying the ac ringing signal. That signal is normally at 20 Hz and measures between 70 and 130 volts rms. It uses that power for the tone generator and to drive the piezoelectric transducer. The sound that is produced is a warble that varies between two frequencies, $f_0/4$ (f_0 − 4) $f_0/5$. The clock, or fundamental, frequency, f_0, is generated by a relaxation oscillator. That oscillator has R2 and C2 as its frequency setting components providing a selectable range of 1 kHz to 10 kHz. Selecting different values for R2 and/or C2 changes the clock frequency, which in turn varies the warble frequencies. The MC34012 chip comes in three different warble rates at which the warble frequencies ($f_0/4$, $f_0/5$) are varied. These warble rates are $f_0/320$, $f_0/640$, or $f_0/160$ and the different chips are designated as MC34012-1, -2, and -3, respectively. For example: with a 4.40 kHz oscillator frequency, the MC34012-1 produces 800 Hz and 1000 Hz tones with a 12.5 Hz warble rate. The MC34012-2 generates 1600 Hz and 2000 Hz tones with a similar 12.5 Hz warble frequency from an 8.0 kHz oscillator frequency. MC34012-3 will produce 400 Hz and 500 Hz tones with a 12.5 warble rate from a 2.0 kHz oscillator frequency.

TONE RINGER II





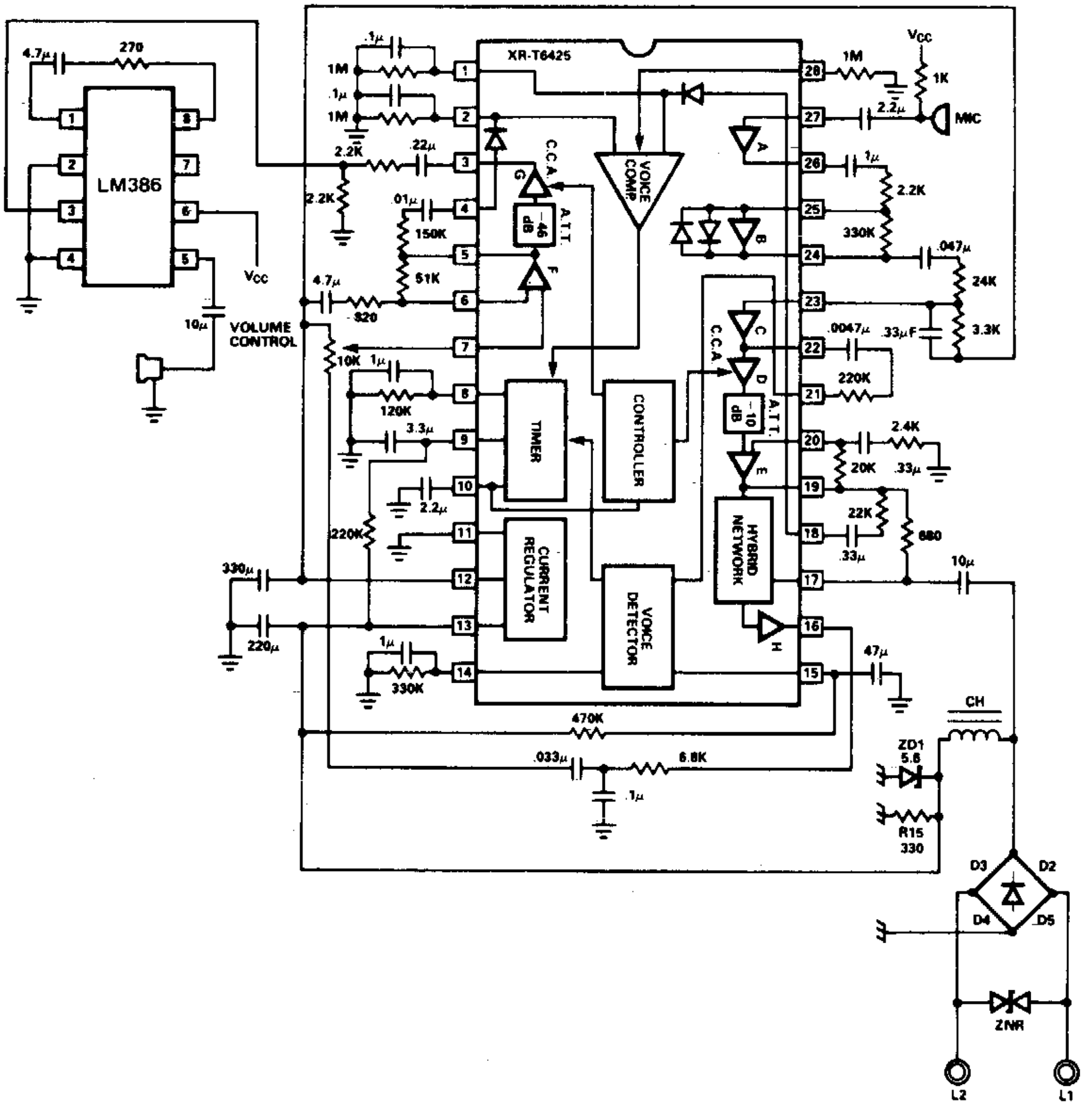
EXAR

Fig. 98-17

Circuit Notes

The XR-T8205 Tone Ringer is primarily intended as a replacement for the mechanical telephone bell. The device can be powered directly from telephone ac ringing voltage or from a separate dc supply. An adjustable trigger level is provided with an external resistor. The circuit is designed for nominal 15 volt operation.

SPEAKERPHONE



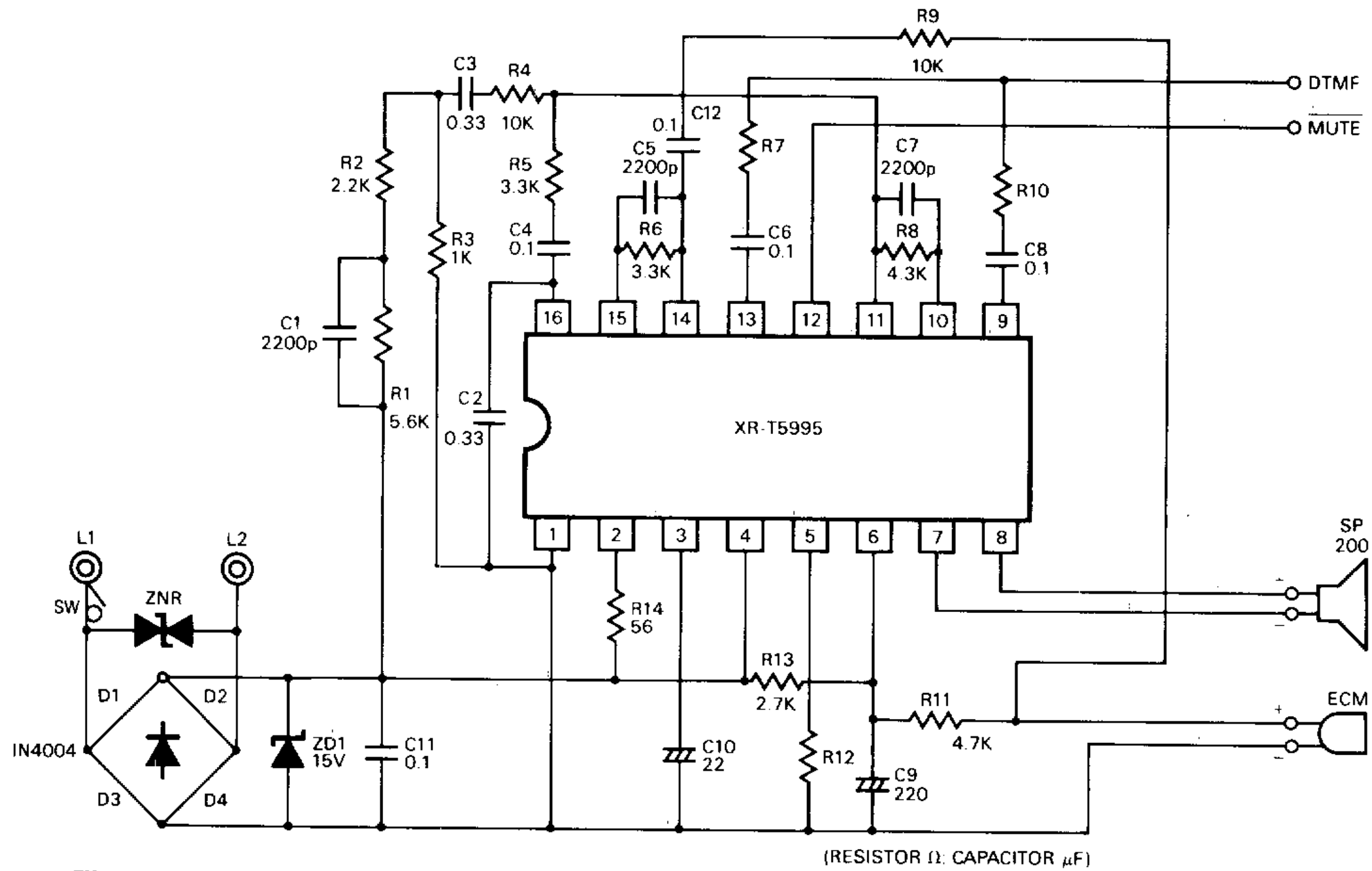
EXAR

Fig. 98-18

Circuit Notes

The XR-T6425 Speakerphone IC makes it possible to carry on conversation without using the handset, while the user is talking into a microphone and listening from a loudspeaker. It is ideal for hands-free conference calls. The XR-T6425 contains most of the circuits to eliminate singing and excessive background noise.

SPEECH NETWORK



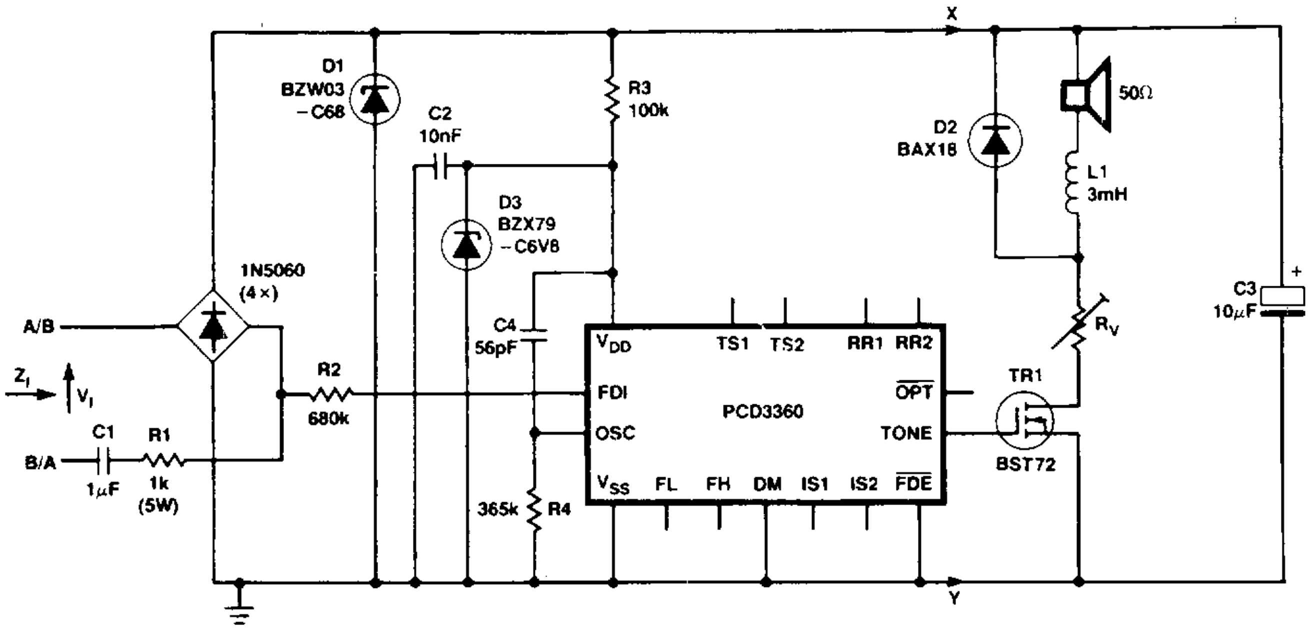
EXAR

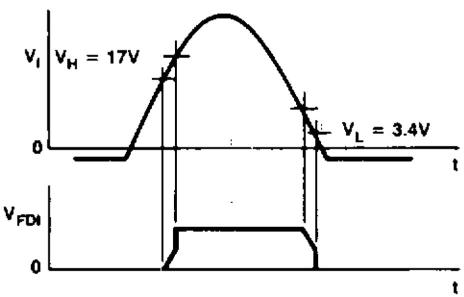
Fig. 98-19

Circuit Notes

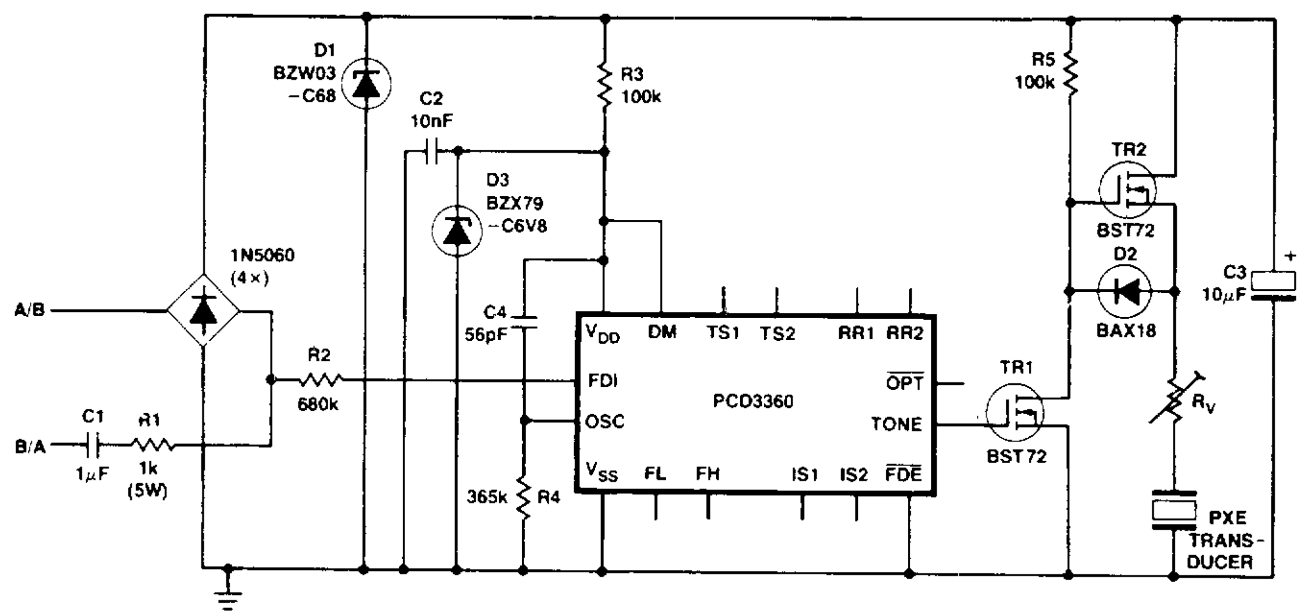
The XR-T5995 Speech Network is a monolithic integrated circuit specifically designed for implementing a low cost telephone circuit. It is designed to use an electrodynamic microphone and electromagnetic receiver to replace a carbon microphone and telephone network hybrid.

PROGRAMMABLE MULTI-TONE TELEPHONE RINGER





Transformerless Electronic Ringer With PCD3360 and a Loudspeaker



PCD3360 Ringer With PXE Transducer

SIGNETICS

Fig. 98-20

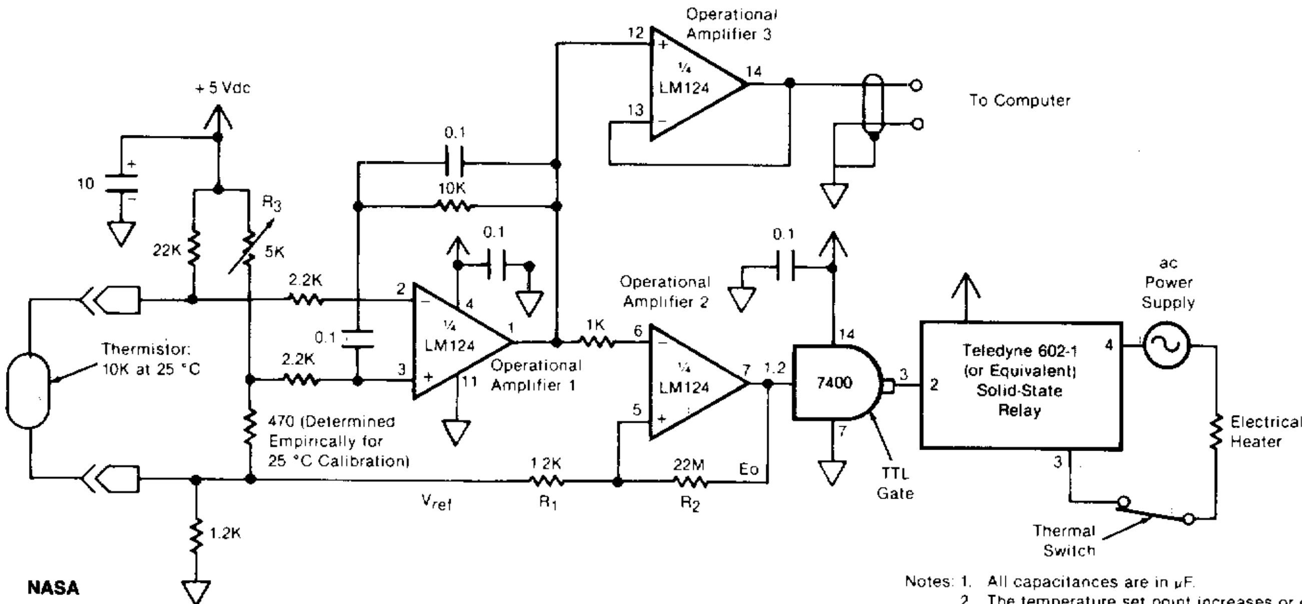
Circuit Notes

Two BST72 transistors provide an output voltage swing almost equal to the voltage at C3. Pins IS1 and IS2 are inoperative because DM = HIGH. Volume control is possible using resistor R_V.

99
Temperature Controls

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TEMPERATURE-CONTROLLING CIRCUIT



NASA

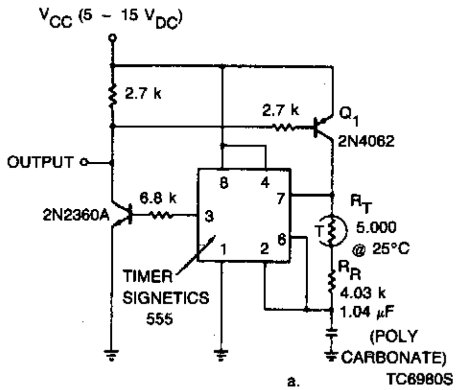
Fig. 99-1

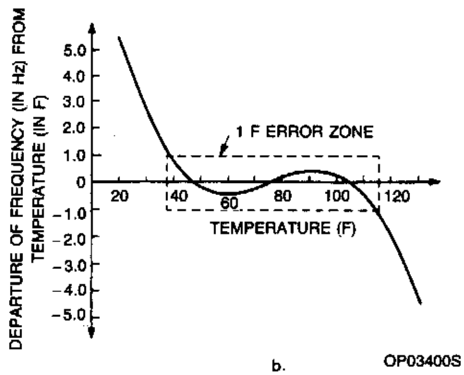
Notes:
1. All capacitances are in μF.
2. The temperature set point increases or decreases with R_3, which is set at 2.17K for 25 °C.
3. The upper and lower tripping voltages of operational amplifier 2 are given by $V_{trip} = V_{ref} \pm E_0 R_1/(R_1 + R_2)$, where V_{ref} = the reference voltage and E_0 = the output voltage of the operational amplifier

Circuit Notes

The circuit switches the current to an electrical heater on and off to maintain the temperature of a room at 25 ±0.5°C. The temperature sensor is a thermistor which provides a differential input (for reduced noise) to an operational amplifier. A 5 kilohm potentiometer is used to adjust the set point through a voltage divider; a value of 2.17 kilohms yields the 25°C setting. A second operational amplifier is connected as an inverting differential-input comparator. The output of operational amplifier 2 controls the electrical heater through a zero-crossing solid-state relay. A transistor/transistor-logic (TTL) gate adjusts the output to the proper level for the relay. A thermal switch is placed in series with the heater and the ac supply for safety in case of thermal runaway. A third operational amplifier monitors the output of the thermistor, providing a signal to a computer for data logging.

TEMPERATURE CONTROL





NOTE:
All resistor values are in ohms.

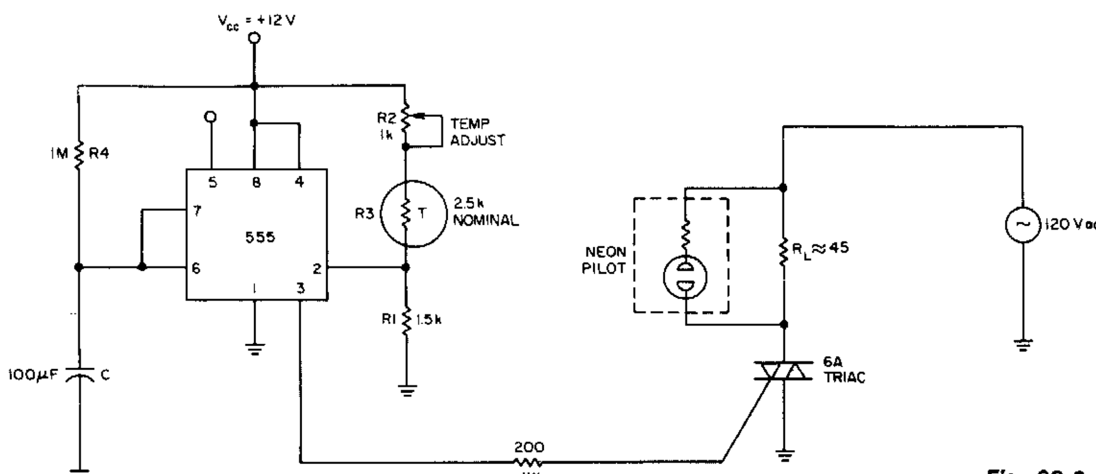
SIGNETICS

Fig. 99-2

Circuit Notes

A couple of transistors and a thermistor in the charging network of the 555 type timer enable this device to sense temperature and produce a corresponding frequency output. The circuit is accurate to within ±1 Hz over a 78°F temperature range.

LOW-COST TEMPERATURE CONTROLLER



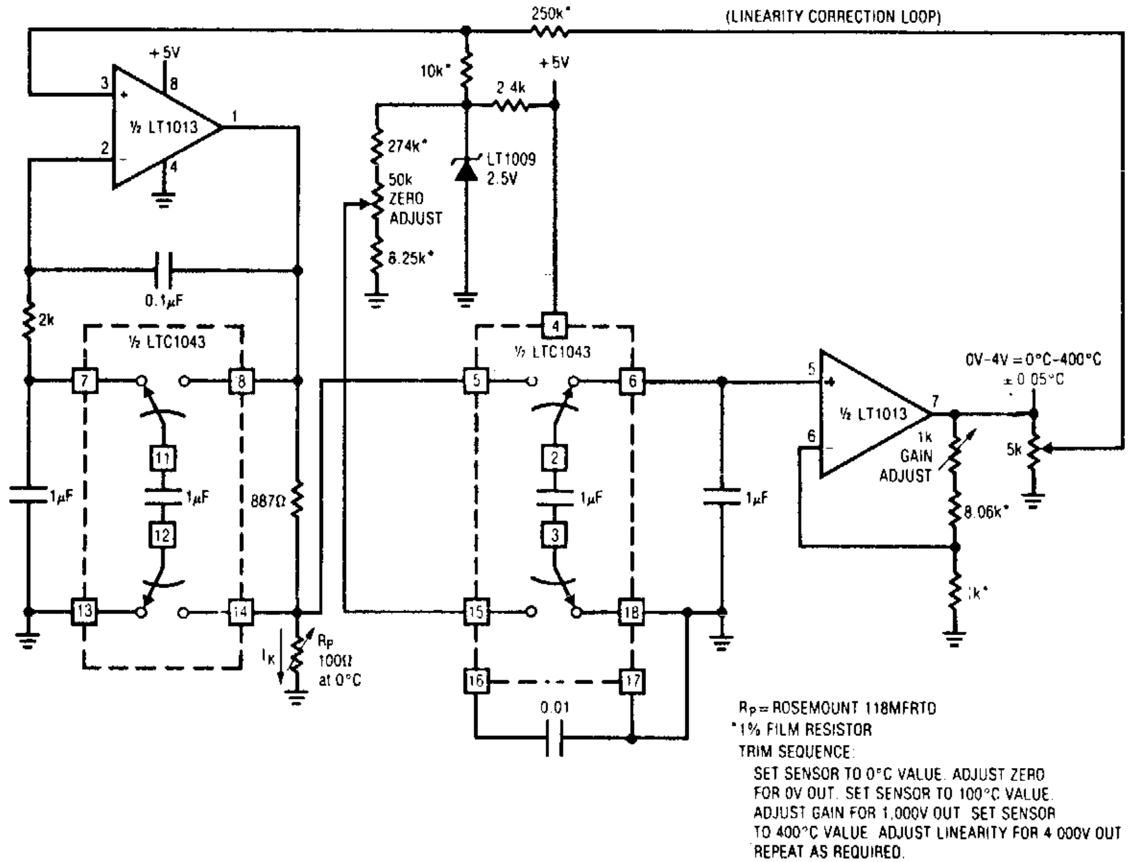
ELECTRONIC DESIGN

Fig. 99-3

Circuit Notes

The internal comparator of the 555 timer, combined with a thermistor, makes a low-cost temperature controller. Resistor R2 sets the temperature trip point.

PRECISION, LINEARIZED PLATINUM RTD SIGNAL CONDITIONER

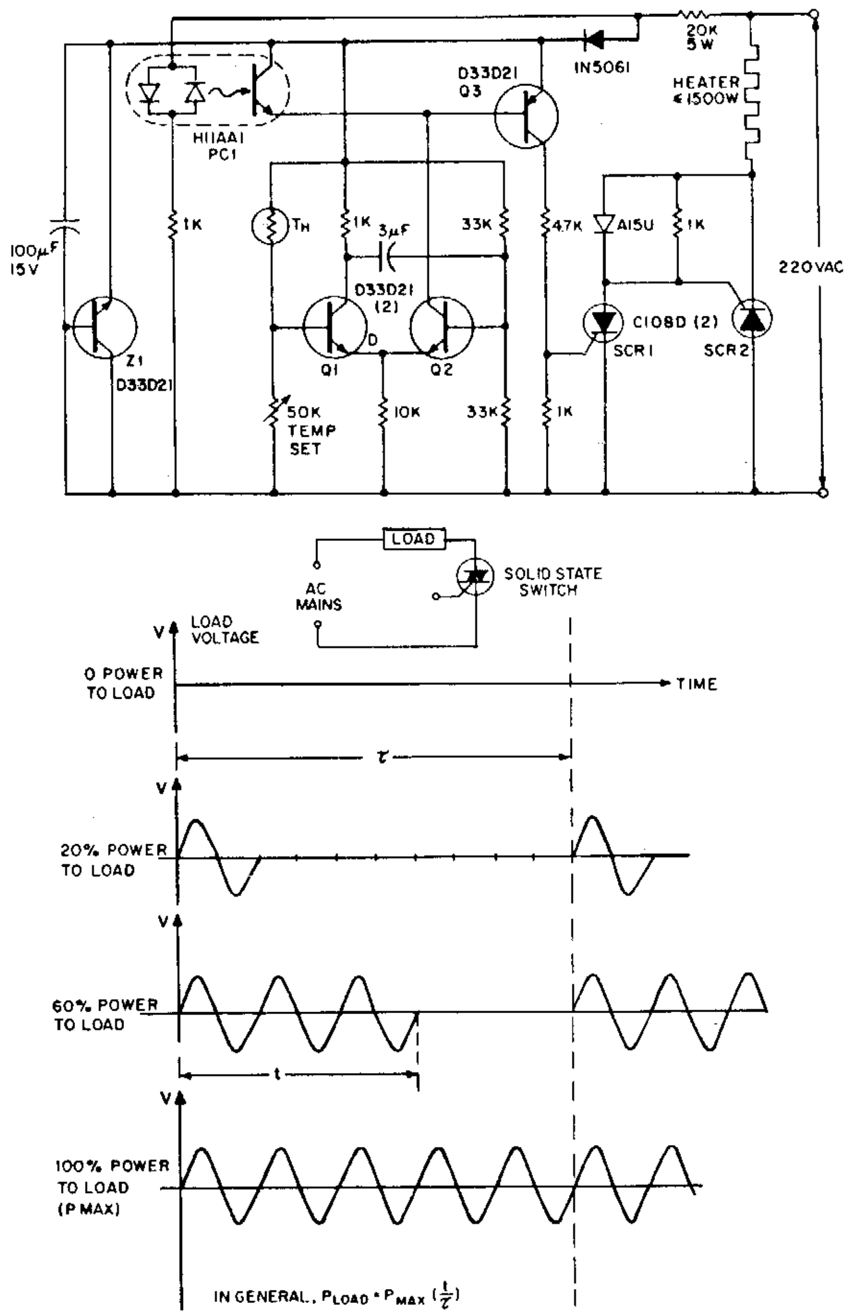


LINEAR TECHNOLOGY CORPORATION ***Fig. 99-4***

Circuit Notes

The circuit provides complete, linearized signal conditioning for a platinum RTD. This LTC1043 based circuit is considerably simpler than instrumentation or multi-amplifier based designs, and will operate form a single 5 V supply. A1 serves as a voltage-controlled ground referred current source by differentially sensing the voltage across the 998 phm feedback resistor. The LTC1043 section which does this presents a single-ended signal to A1's negative input, closing a loop. The 2 k 0.1 μF combination sets amplifier roll-off well below the LTC1043's switching frequency and the configuration is stable. Because A1's loop forces a fixed voltage across the 887 ohm resistor, the current through R_P is constant. A1's operating point is primarily fixed by the 2.5 V LT1009 voltage reference.

LOW POWER ZERO VOLTAGE SWITCH TEMPERATURE CONTROLLER



GENERAL ELECTRIC

Fig. 99-5

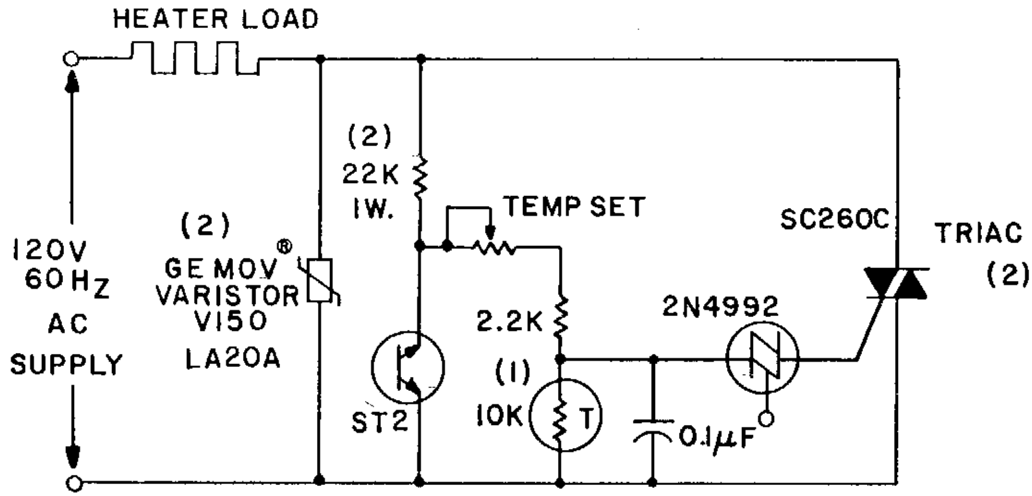
Circuit Notes

The "zero voltage switching" technique is widely used to modulate heating and similar types of ac loads where the time constant associated with the load (tens of seconds to minutes) is sufficiently long to allow smooth proportional modulation by time ratio control, using one complete cycle of the ac input voltage as the minimum switching movement. Despite its attractions, the traditional triac-based ZVS is virtually unusable for the control of very low power loads, especially from 220 volt ac inputs due to the triac's reluctance to latch-on into the near-zero instantaneous currents that flow through it and the load near the ac voltage zero crossover points. The circuit side-steps the latching problem by employing a pair of very sensitive low current reverse blocking thyristors (C106) connected in antiparallel; these are triggered by a simple thermistor modulated differential amplifier (Q1, Q2), with zero voltage logic furnished by an H11AA1 ac input optocoupler. With the NTC thermistor TH calling for heat, transistor Q1 is cut off and Q2 is on, which would normally provide continuous base drive to Q3, with consequent triggering of either SCR, or of SCR 2 via SCR1, depending on phasing of the ac input.

Note that when the ac input voltage is positive with respect to SCR 2, SCR 1 is reverse biased and, in the presence of "gate" current from Q3, behaves as a remote base transistor, whose output provides via blocking diode CR1, positive gate trigger current for SCR 2. When the ac input polarity is reversed (SCR 1's anode positive), SCR 1 behaves as a direct fired conventional thyristor. "Trigger" current to SCR 1, however, is not continuous, even when TH is calling for heat and Q2 is delivering base current to Q3. In this situation, Q3 is inhibited from conduction by the clamping action of PC1, an H11AA photocoupler, except during those brief instants when the ac input voltage is near zero and the coupler input diodes are deprived of current.

Triggering of either SCR can occur only at ac voltage crossing points, and RFI-less operation results. The proportional control feature is injected via the positive feedback action of capacitor CM, which converts the differential amplifier Q1, Q2 into a simple multivibrator, whose duty cycle varies from one to 99 percent according to the resistance of TH. Zener diode Z1 is operational, being preferred when maximum immunity from ac voltage induced temperature drift is desired.

HEATER ELEMENT TEMPERATURE CONTROLLER



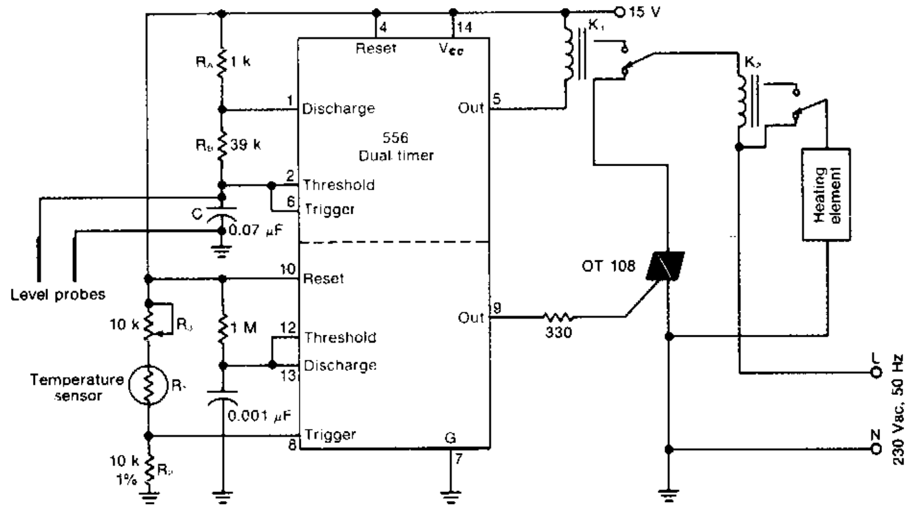
NOTES:
1. Thermistor National Lead type 1D101, or equivalent.
2. Component values for 220V operation:
   Resistor – 47K, 2W
   GE-MOV® Varistor – V275LA20A
   Triac – SC260E

GENERAL ELECTRIC *Fig. 99-6*

Circuit Notes

The circuit can control up to 6 kW of heating, with moderate gain, using a 25-amp triac (SC260D). Feedback is provided by the negative temperature co-efficient (NTC) thermistor, which is mounted adjacent to the environment being temperature controlled. The temperature set potentiometer is initially adjusted to the desired heating level. As the thermistor becomes heated by the load, its resistance drops, phasing back the conduction angle of the triac, so the load voltage is reduced. The ST2 diac is used as a back-to-back zener diode. Its negative resistance region in its E-I characteristic provides a degree of line voltage stabilization. As the input line voltage increases, the diac triggers earlier in the cycle and, hence, the average charging voltage to the 0.1 μF capacitor, decreases.

DUAL-TIMER CHIP CONTROLS TEMPERATURE WHILE MONITORING LIQUID LEVEL



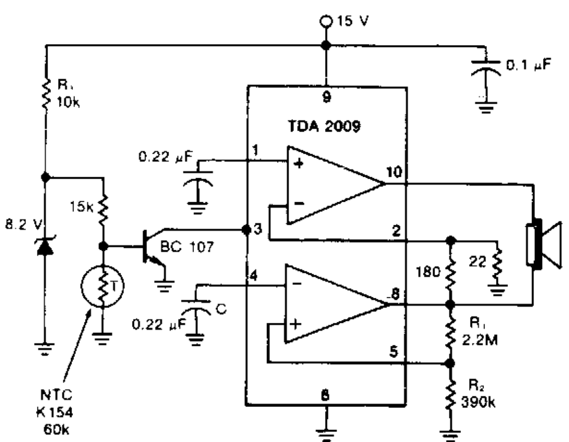
ELECTRONIC DESIGN

Fig. 99-7

Circuit Notes

One-half of a 556 dual timer monitors the temperature of a liquid bath, controlling a heating element that maintains temperature within ±2°C over a 32° - 200°C range. The other half monitors the liquid level, disconnecting the heater when the level drops below a preset point.

TEMPERATURE ALARM



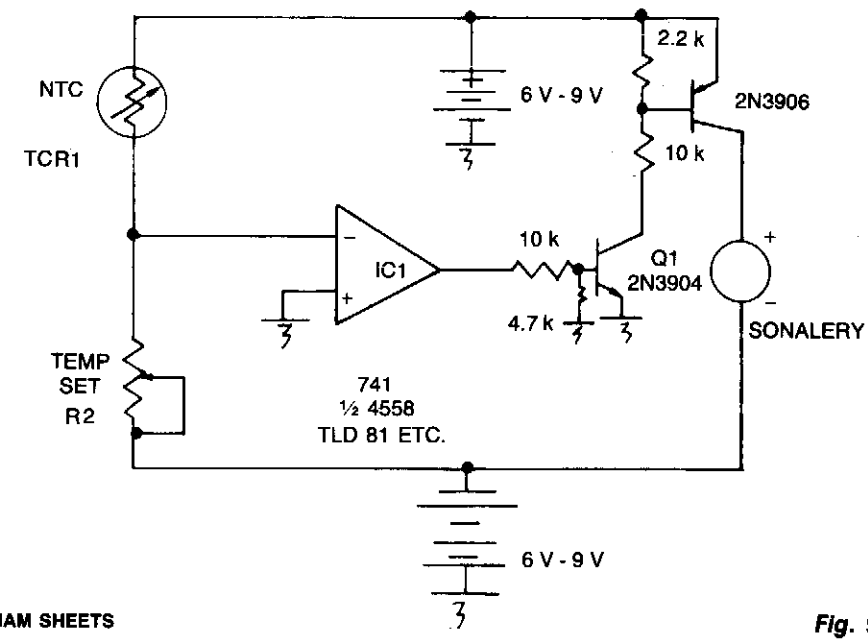
ELECTRONIC DESIGN

Fig. 99-8

Circuit Notes

The mute pin of this dual audio amplifier is used as the trigger for a one chip high-temperature alarm. One-half of the IC is connected as an oscillator and the other boosts the audio alarm outputs to 10W.

ADJUSTABLE THRESHOLD TEMPERATURE ALARM



WILLIAM SHEETS

Fig. 99-9

Circuit Notes

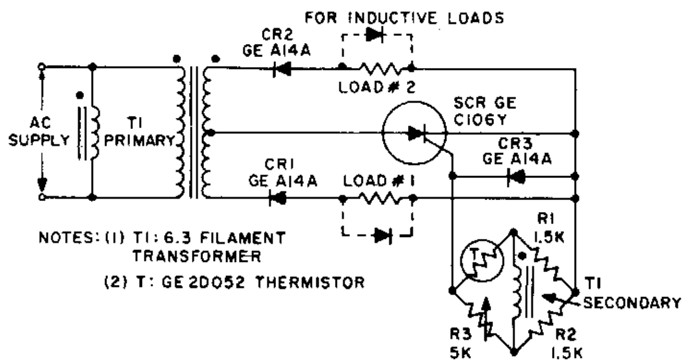
When R1 increases as temp decreases, the output of IC1 goes positive, turning on Q1. Q1 conducts and causes Q2 to conduct, turning on the audible alarm. The threshold is set with potentiometer R2.

100

Temperature Sensors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DUAL OUTPUT, OVER-UNDER TEMPERATURE MONITOR



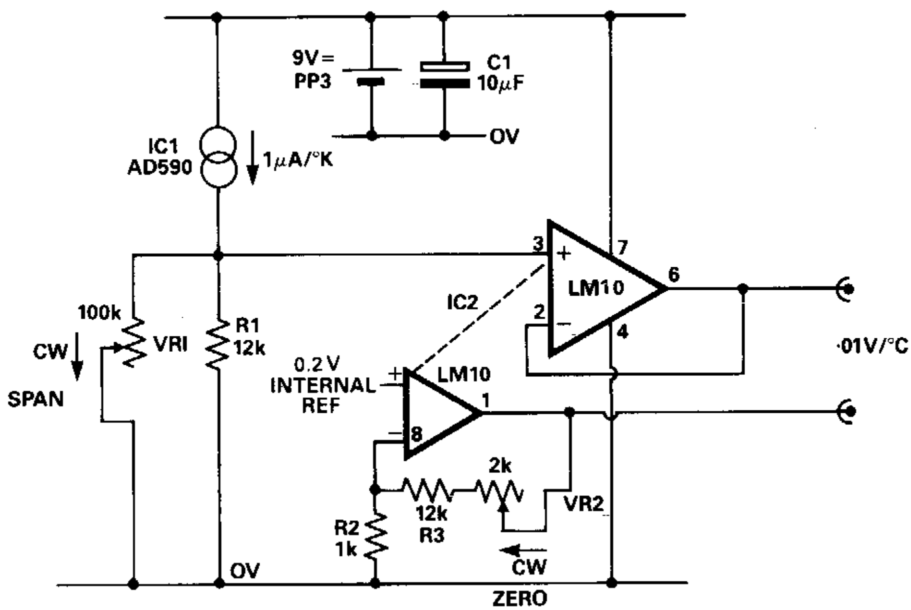
GENERAL ELECTRIC

Fig. 100-1

Circuit Notes

This circuit is ideal for use as an over-under temperature monitor, where its dual output feature can be used to drive HIGH and LOW temperature indicator lamps, relays, etc. T1 is a 6.3 volt filament transformer whose secondary winding is connected inside a four arm bridge. When the bridge is balanced, ac output is zero, and C5 (or C7) receives no gate signal. If the bridge is unbalanced by raising or lowering the thermistor's ambient temperature, and ac voltage will appear across the SCR's gate cathode terminals. Depending in which sense the bridge is unbalanced, the positive gate voltage will be in phase with, or 180° out of phase with the ac supply. If the positive gate voltage is in phase, the SCR will deliver load current through diode CR1 to load (1), diode CR2 blocking current to load (2). Conversely, if positive gate voltage is 180° out of phase, diode CR2 will conduct and deliver power to load (2), CR1 being reverse biased under these conditions. With the component values shown, the circuit will respond to changes in temperature of approximately 1-2°C. Substitution of other variable-resistance sensors, such as cadmium sulfide light dependent resistors (LDR) or strain gauge elements, for the thermistor shown is permissible.

TEMPERATURE SENSOR AND DVM INTERFACE



ELECTRONIC ENGINEERING

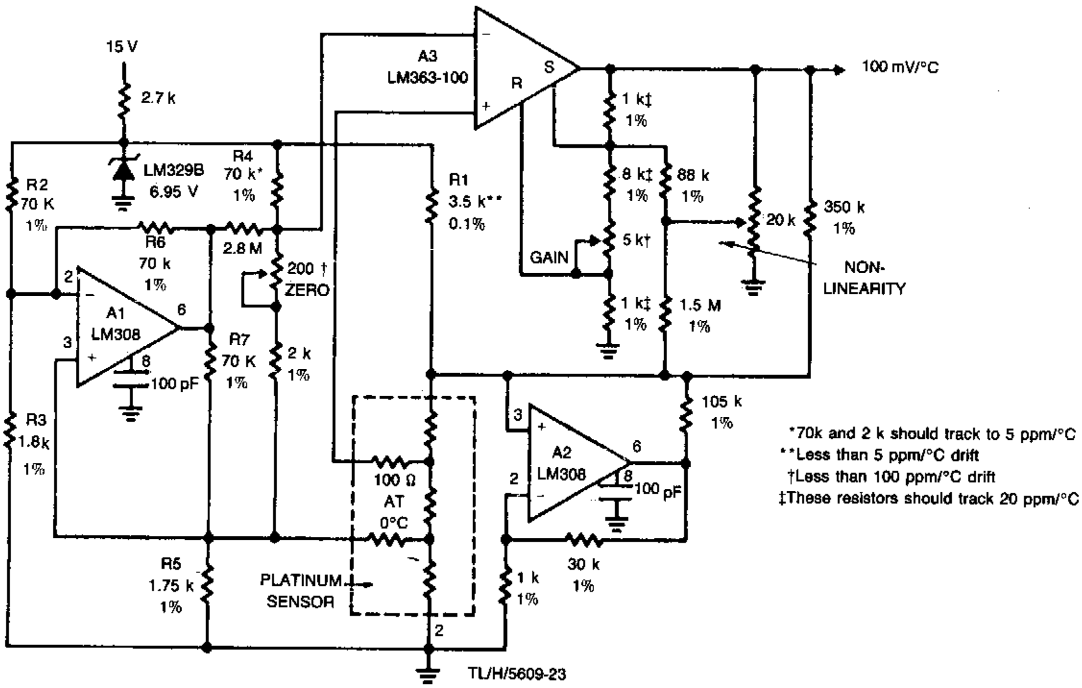
Fig. 100-2

Circuit Notes

The DVM gives a direct indication of the temperature of the sensor in degrees Centigrade. The temperature sensor IC1 gives a nominal 1 μA per degree Kelvin which is converted to 10 mV per degree Kelvin by R1 and VR1. IC2 is a micropower, low input drift op amp with internal voltage reference and amplifier. The main op amp in IC1 is connected as a voltage follower to buffer the sensor voltage at R1.

The second amplifier in IC1 is used to amplify the .2 V internal reference up to 2.73 V in order to offset the 273 degrees below 0°C. The output voltage of the unit is the differential output of the two op amps and is thus equal to 0.01 V per °C.

CURVATURE CORRECTED PLATINUM RTD THERMOMETER

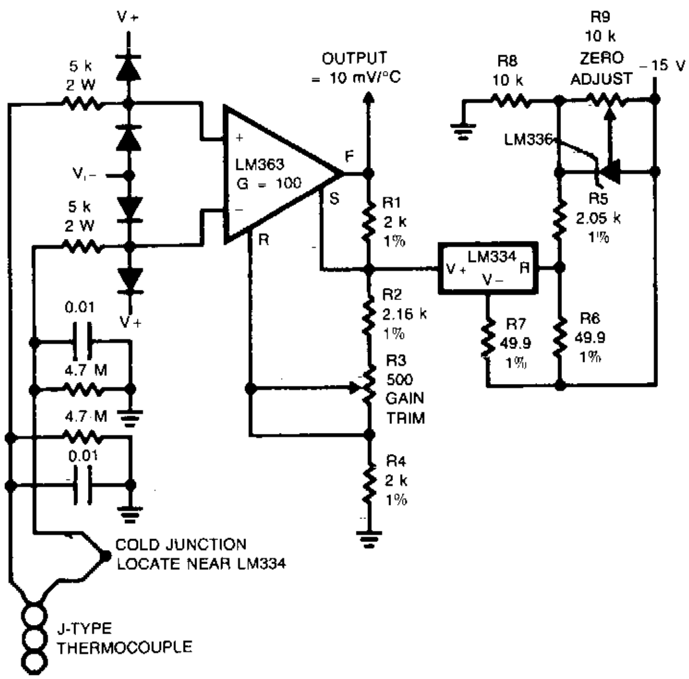


NATIONAL SEMICONDUCTOR CORP. *Fig. 100-3*

Circuit Notes

This thermometer is capable of 0.01°C accuracy over −50°C to +150°C. A unique trim arrangement eliminates cumbersome trim interactions so that zero gain, and nonlinearity correction can be trimmed in one even trip. Extra op amps provide full Kelvin sensing on the sensor without adding drift and offset terms found in other designs. A1 is configured as a Howland current pump, biasing the sensor with a fixed current. Resistors R2, R3, R4 and R5 form a bridge driven into balance by A1. In balance, both inputs of A2 are at the same voltage. Since R6 = R7, A1 draws equal currents from both legs of the bridge. Any loading of the R4/R5 leg by the sensor would unbalance the bridge; therefore, both bridge taps are given to the sensor open circuit voltage and no current is drawn.

THERMOCOUPLE AMPLIFIER WITH COLD JUNCTION COMPENSATION



NATIONAL SEMICONDUCTOR CORP. ***Fig. 100-4***

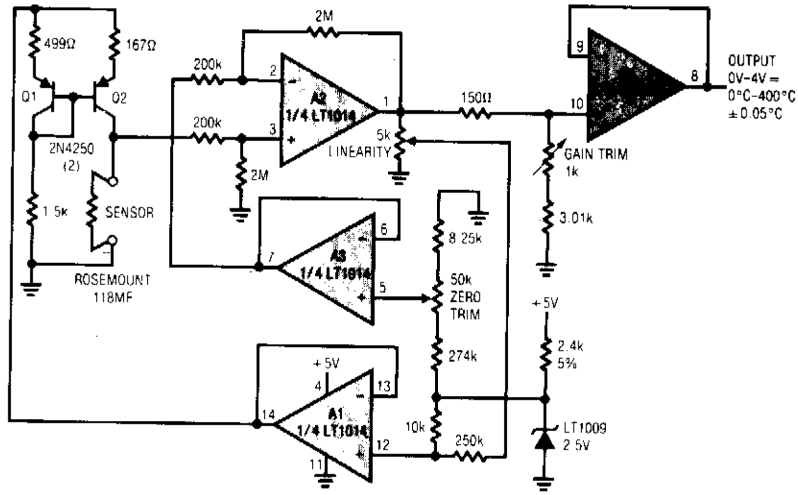
Circuit Notes

Input protection circuitry allows thermocouple to short to 120 Vac without damaging the amplifier.

Calibration:

1. Apply a 50 mV signal in place of the thermocouple. Trim R3 for $V_{OUT} = 12.25$ V.
2. Reconnect the thermocouple. Trim R9 for correct output.

5-V POWERED, LINEARIZED PLATINUM RTD SIGNAL CONDITIONER

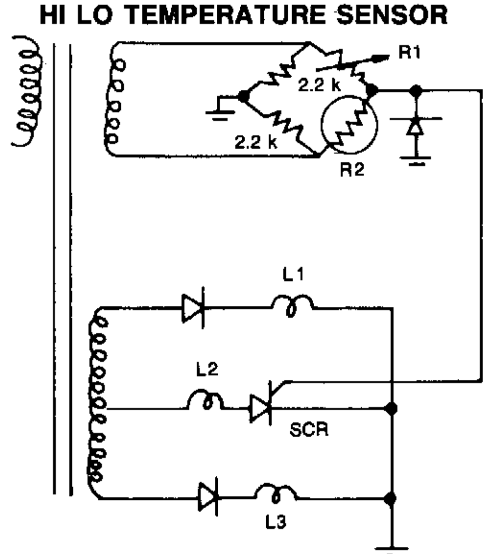


LINEAR TECHNOLOGY CORP.

ALL RESISTORS ARE TRW-MAR-6 METAL FILM.
RATIO MATCH 2M 200K ±0.01%.
TRIM SEQUENCE
SET SENSOR TO 0 VALUE
ADJUST ZERO FOR 0V OUT

Fig. 100-5

HI LO TEMPERATURE SENSOR



WILLIAM SHEETS

Fig. 100-6

Circuit Notes

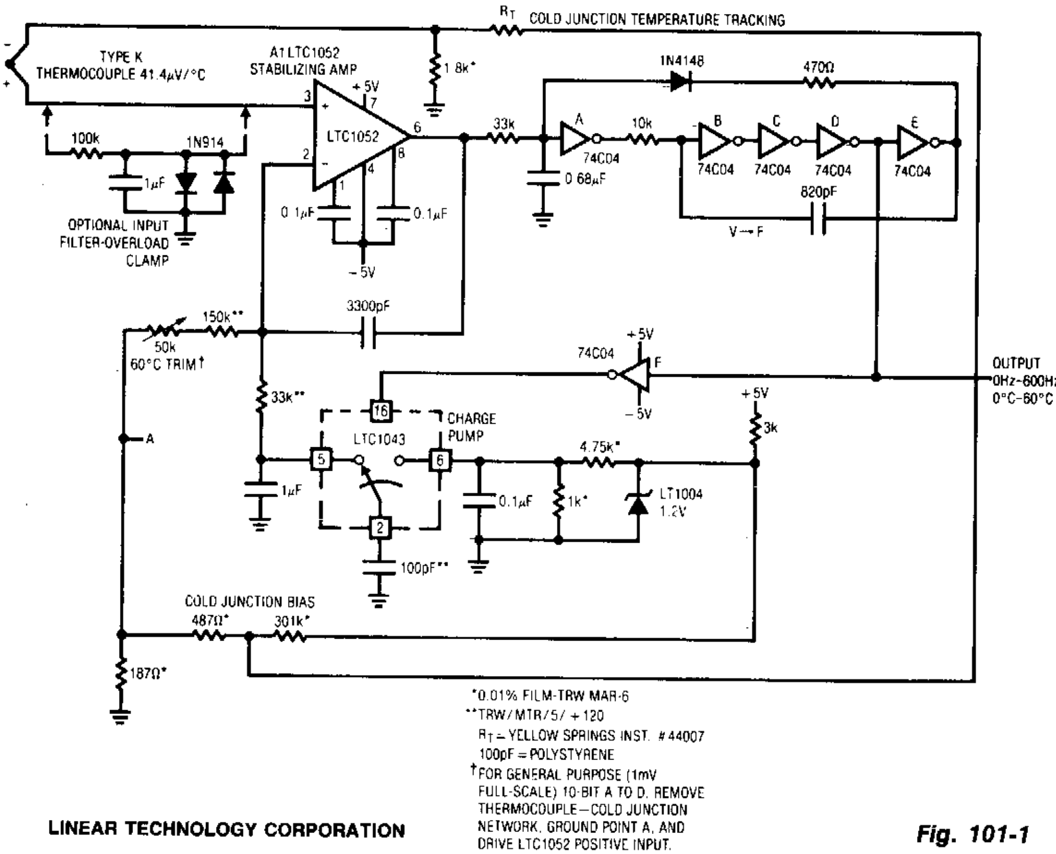
Resistors R1, R2, and the two 2.2 k resistors form a bridge circuit. R2 is a thermistor, and R1 sets the temperature at which L2 lights. Lower or higher temperatures light L1 or L3 to indicate an over- or under-temperature condition.

101

Temperature-to-Frequency Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TEMPERATURE-TO-FREQUENCY CONVERTER

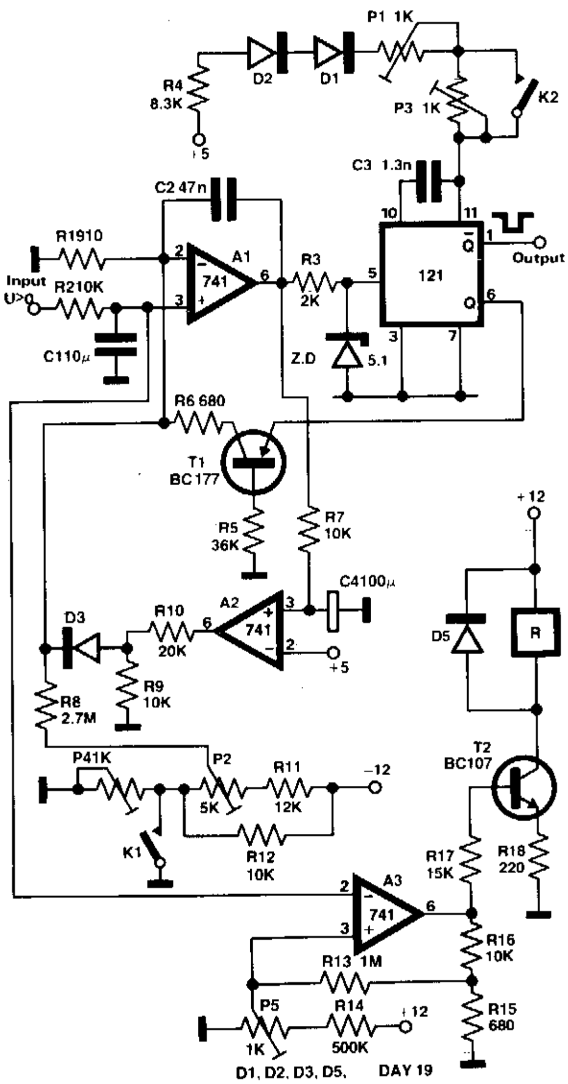


LINEAR TECHNOLOGY CORPORATION ***Fig. 101-1***

Circuit Notes

A1's positive input is biased by the thermocouple. A1's output drives a crude V → F converter, comprised of the 74C04 inverters and associated components. Each V → F output pulse causes a fixed quantity of charge to be dispensed into the 1 μF capacitor from the 100 pF capacitor via the LT1043 switch. The larger capacitor integrates the packets of charge, producing a dc voltage at A1's negative input. A1's output forces the V→ F converter to run at whatever frequency is required to balance the amplifier's inputs. This feedback action eliminates drift and nonlinearities in the V → F converter as an error item and the output frequency is solely a function of the dc conditions at A1's inputs. The 3300 pF capacitor forms a dominant response pole at A1, stabilizing the loop.

DIGITAL TEMPERATURE MEASURING CIRCUIT



ELECTRONIC ENGINEERING

Fig. 101-2

Circuit Notes

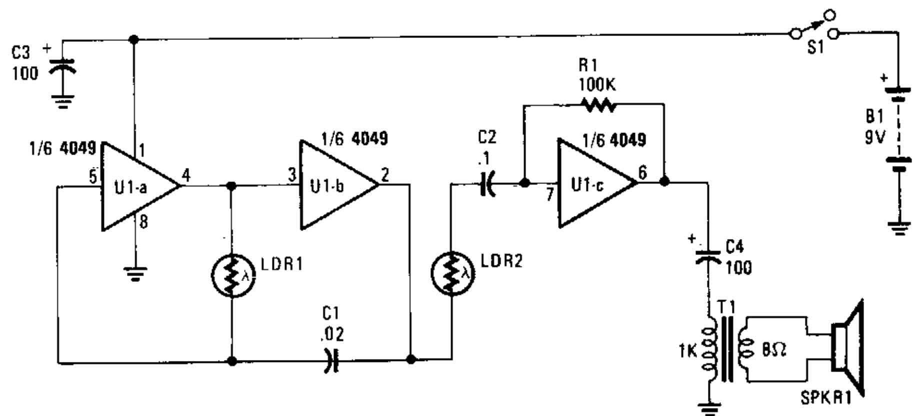
The output voltage of a thermocouple is converted into frequency measured by a digital frequency meter. The measuring set connected with Ni-NiCr thermocouple permits you to measure the temperatures within the range of 5°C - 800°C with ±1°C error. The output thermocouple signal is proportional to the temperature difference between the hot junction and the thermostat kept at 0°C, it drives the voltage-to-frequency converter changing the analogue input signal into the output frequency with the conversion ratio adjusted in such a way, that the frequency is equal to the measured temperature in Celsius degrees, e.g., for 350°C the frequency value is 350 Hz.

102

Theremins

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

ELECTRONIC THEREMIN



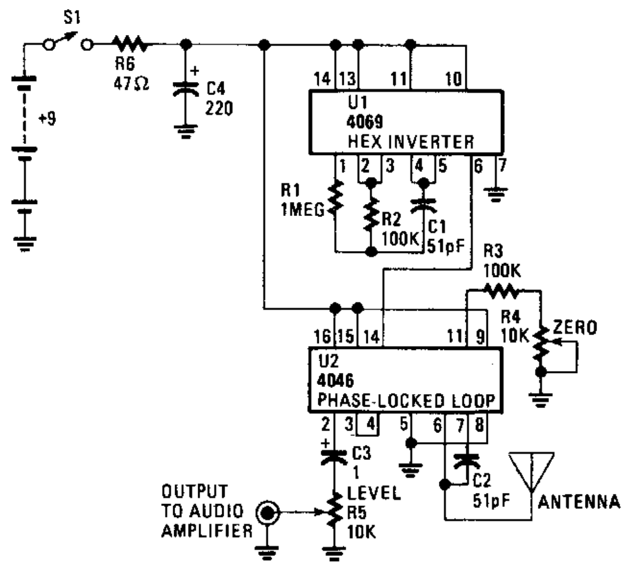
HANDS-ON ELECTRONICS

Fig. 102-1

Circuit Notes

This circuit has the CMOS IC doing double-duty performance. The first two inverters operate as a digital audio oscillator; the third operates as a low-gain linear audio amplifier. As the intensity of the light falling on photoresistor LDR1 increases the oscillator's frequency increases; similarly, the illumination falling on photoresistor LDR2 determines the volume level from the loudspeaker: The more illumination the more volume. If you flop and wave your hands between the two photocells and a light source, a special kind of electronic music will be produced.

DIGITAL THEREMIN



HANDS-ON ELECTRONICS *Fig. 102-2*

Circuit Notes

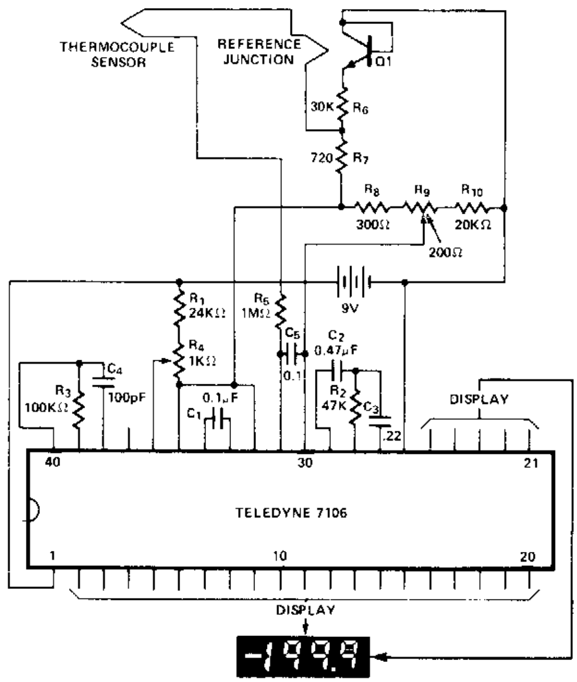
The CD4069 or 74C04 hex inverter—is used as a fixed-frequency oscillator centered around 100 kHz. U2 contains the variable frequency oscillator and balanced modulator. The CD4046 is a phase-locked loop and R3, R4, and C2 determine the center frequency of the on-chip oscillator. The antenna forms a parallel capacitance with C2, which allows the frequency to be shifted several kilohertz by bringing a hand near the antenna. R4, the ZERO control, allows the variable oscillator to be set to the same frequency as the fixed oscillator. When the difference frequency is below 15 Hz, it is below the lower frequency limit of the ear. By setting both oscillators to the same frequency, the Theremin remains silent until the performer brings his or her hand near the antenna. The oscillators are mixed by an exclusive OR gate inside the 4046. That gate acts as a digital balanced modulator, which produces the sum and difference frequencies. The output of the gate is then ac coupled by C3 to LEVEL control R5 and an output jack for connection to an audio amplifier or stereo receiver.

103

Thermometer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DIGITAL THERMOCOUPLE THERMOMETER



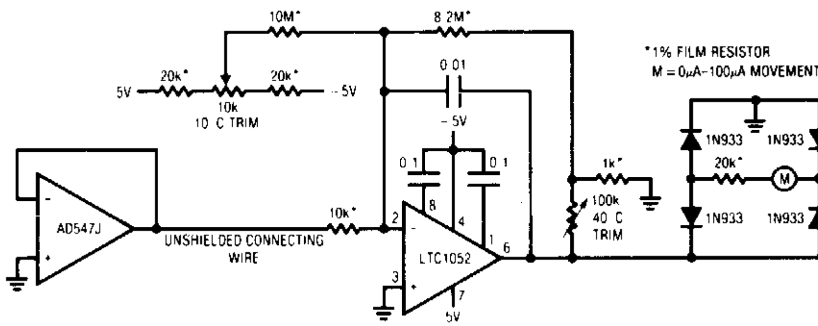
TELEDYNE SEMICONDUCTOR

Fig. 103-1

Circuit Notes

This digital thermocouple thermometer uses one active component and 15 passive components. With this circuit, both type J and type K thermocouples may be used. The type J will measure over the temperature range of 10 to 530°C with a conformity of ±2°C. The type K will measure over a temperature range of 0°C to 1000°C with a conformity of ±3°C.

REMOTE THERMOMETER

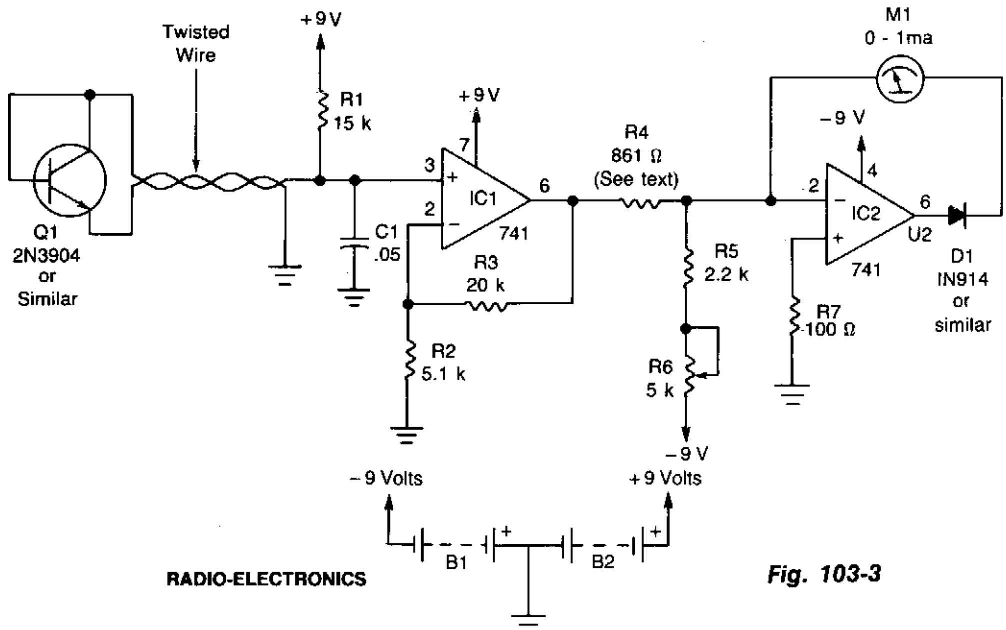


LINEAR TECHNOLOGY ***Fig. 103-2***

Circuit Notes

The low output impedance of a closed loop op amp gives ideal line-noise immunity, while the op amp's offset voltage drift provides a temperature sensor. Using the op amp in this way requires no external components and has the additional advantages of a hermetic package and unit-to-unit mechanical uniformity if replacement is ever required. The op amp's offset drift is amplified to drive the meter by the LTC1052. The diode bridge connection allows either positive or negative op amp temperature sensor offsets to interface directly with the circuit. In this case, the circuit is arranged for a +10°C to +40°C output, although other ranges are easily accommodated. To calibrate this circuit, subject the op amp sensor to a +10°C environment and adjust the 10°C trim for an appropriate meter indication. Next, place the op amp sensor in a +40°C environment and trim the 40°C adjustment for the proper reading. Repeat this procedure until both points are fixed. Once calibrated, this circuit will typically provide accuracy within ±2°C, even in high noise environments.

ELECTRONIC THERMOMETER



RADIO-ELECTRONICS

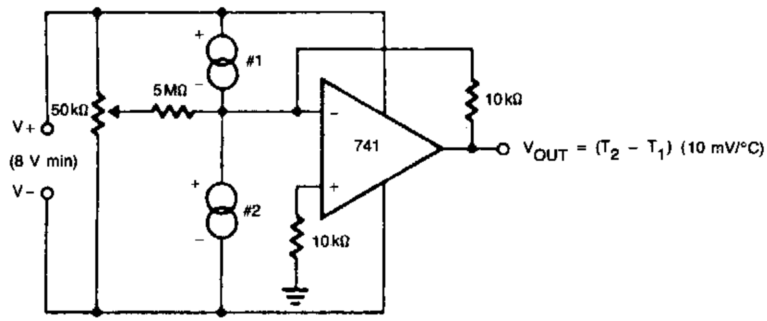
Fig. 103-3

Circuit Notes

An inexpensive electronic thermometer is capable of measuring temperatures over a range of from −30°F to +120°F. A diode-connected 2N3904 transistor used as the temperature sensor forms a voltage divider with R1. As temperature increases, the voltage drop across the transistor changes by approximately −1.166 millivolts-per°F. As a result, the current at pin 3 of IC1, a 741 op amp with a gain of 5, decreases as the temperature measured by the sensor increases.

A second 741 op amp, IC2 is configured as an inverting amplifier. Resistors R5 and R6 calibrate the circuit. Calibration is also straightforward. When properly done, a temperature of −30°F will result in a meter reading of 0 milliamps, while a temperature of 120°F will result in a meter reading of 1 milliamp. Divide the scale between those points into equal segments and mark the divisions with the appropriate corresponding temperatures. The calibration is completed by placing the sensor in an environment with a known temperature, such as an ice-point bath. Place the sensor in the bath and adjust R6 until you get the correct meter reading.

DIFFERENTIAL THERMOMETER



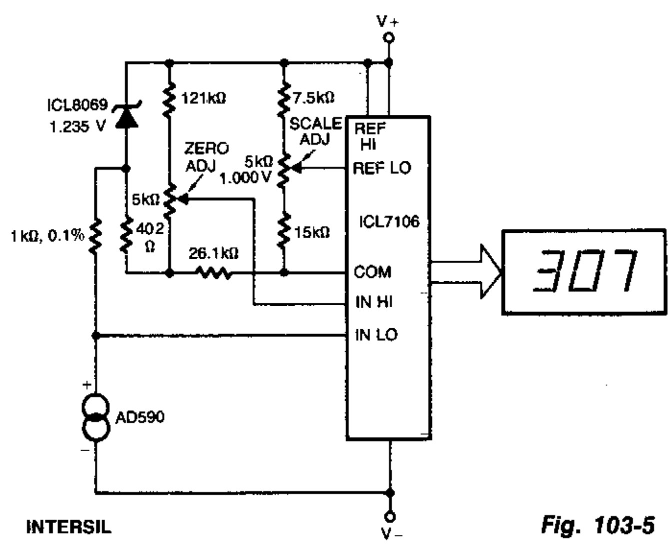
INTERSIL

Fig. 103-4

Circuit Notes

The 50 k ohm pot trims offsets in the devices whether internal or external, so it can be used to set the size of the difference interval. This also makes it useful for liquid-level detection (where there will be a measurable temperature difference).

BASIC DIGITAL THERMOMETER, KELVIN SCALE WITH ZERO ADJUST



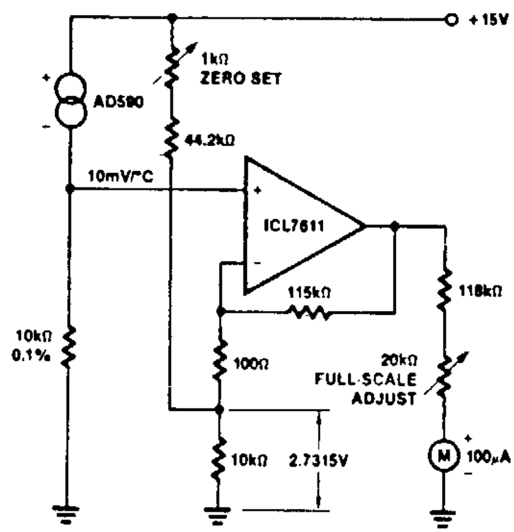
INTERSIL

Fig. 103-5

Circuit Notes

This circuit allows "zero adjustment" as well as slope adjustment. The ICL8069 brings the input within the common-mode range, while the 5 k ohm pots trim any offset at 218° K (−55°C), and set the scale factor.

CENTIGRADE THERMOMETER (0°C-100°C)



INTERSIL

Fig. 103-6

Circuit Notes

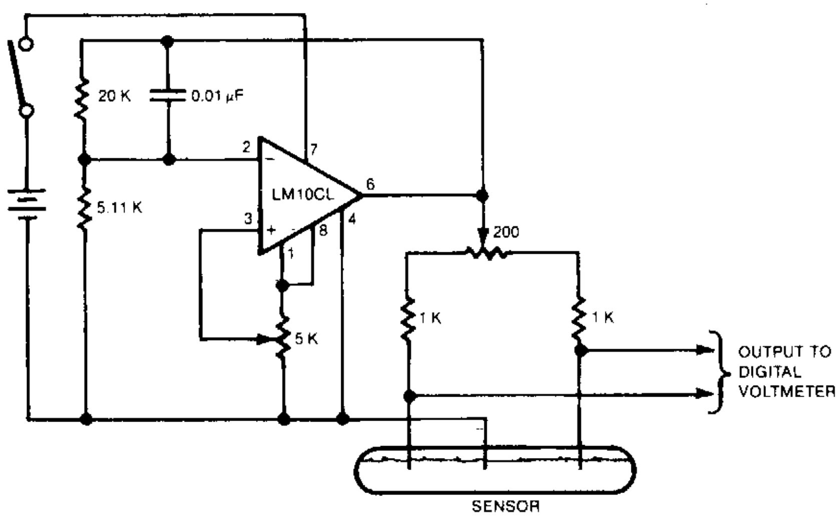
The ultra-low bias current of the ICL7611 allows the use of large-value gain-resistors, keeping meter-current error under ½%, and therefore saving the expense of an extra meter-driving amplifier.

104

Tilt Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TILTMETER INDICATES SENSE OF SLOPE

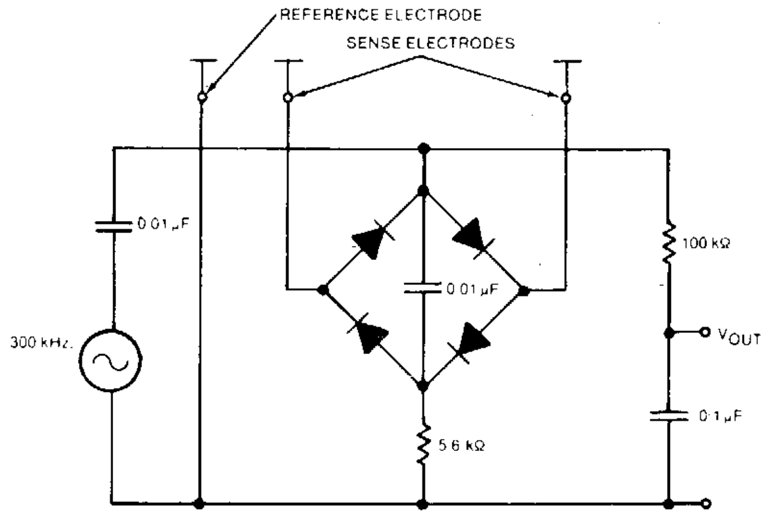


NASA *Fig. 104-1*

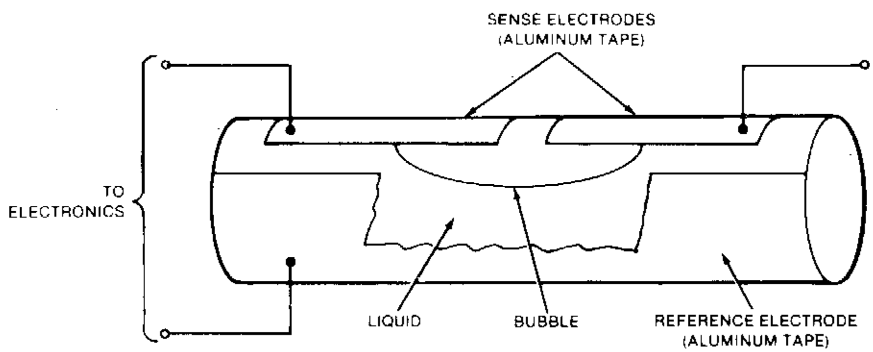
Circuit Notes

Electrodes are immersed in an electrolyte that remains level while the sensor follows the tilt of the body on which it is placed, more of one outer electrode and less of the other are immersed and their resistances fall or rise, respectively. The resistance change causes a change in the output voltage of the bridge circuit. The sensor forms the two lower legs of the bridge, and two 1000 ohm metal film resistors and a 200 ohm ceremet balance potentiometer form the two upper legs. In preparation for use, the bridge is balanced by adjusting the balance potentiometer so that the bridge output voltage is zero when the sensor is level. The bridge input voltage (dc excitation) is adjusted to provide about 10 millivolts output per degree of slope, the polarity indicating the sense of the slope. This scaling factor allows the multimeter to read directly in degrees if the user makes a mental shift of the meter decimal point. The scaling-factor calibration is done at several angles to determine the curve of output voltage versus angle.

DIFFERENTIAL CAPACITANCE MEASUREMENT CIRCUIT



DIFFERENTIAL CAPACITANCE
MEASUREMENT CIRCUIT

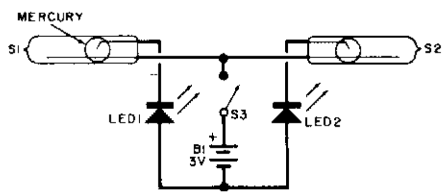


NASA ***Fig. 104-2***

Circuit Notes

A bubble vial with external aluminum-foil electrodes is the sensing element for a simple indicating tiltmeter. To measure bubble displacement, a bridge circuit detects the difference in capacitance between the two sensing electrodes and the reference electrode. Using this circuit, a tiltmeter level vial with 2 mm deflection for 5 arc-seconds of tilt easily resolves 0.05 arc-second. The four diodes are CA3039, or equivalent.

ULTRA-SIMPLE LEVEL



POPULAR ELECTRONICS *Fig. 104-3*

Circuit Notes

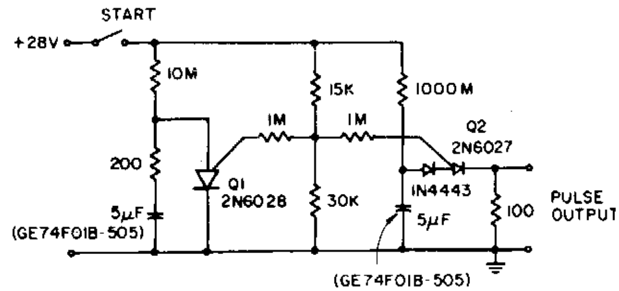
This electronic level uses two LED indicators instead of an air bubble. If the surface is tilted to the right, one LED lights; if it's tilted to the left, the other LED lights. When the surface is level, both LEDs light. It uses two unidirectional mercury switches, S1 and S2. The unidirectional mercury switch has one long electrode and one short, angled electrode. The pool of mercury "rides" on the long electrode and makes contact between the two electrodes if the unit is held in a horizontal position.

105

Time-Delay Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

HOUR TIME-DELAY SAMPLING CIRCUIT



GENERAL ELECTRIC

Fig. 105-1

Circuit Notes

The circuit lowers the effective peak current of the output PUT, Q2. By allowing the capacitor to charge with high gate voltage and periodically lowering gate voltage, when Q1 fires, the timing resistor can be a value which supplies a much lower current than I_P. The triggering requirement here is that minimum charge to trigger flow through the timing resistor during the period of the Q1 oscillator. This is not capacitor size dependent, only capacitor leakage and stability dependent.

TIME DELAY WITH CONSTANT CURRENT CHARGING

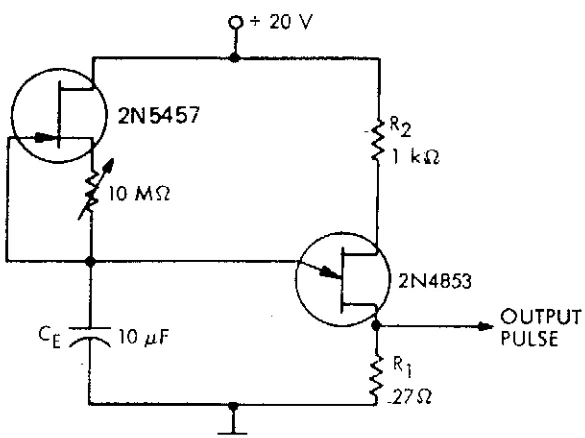
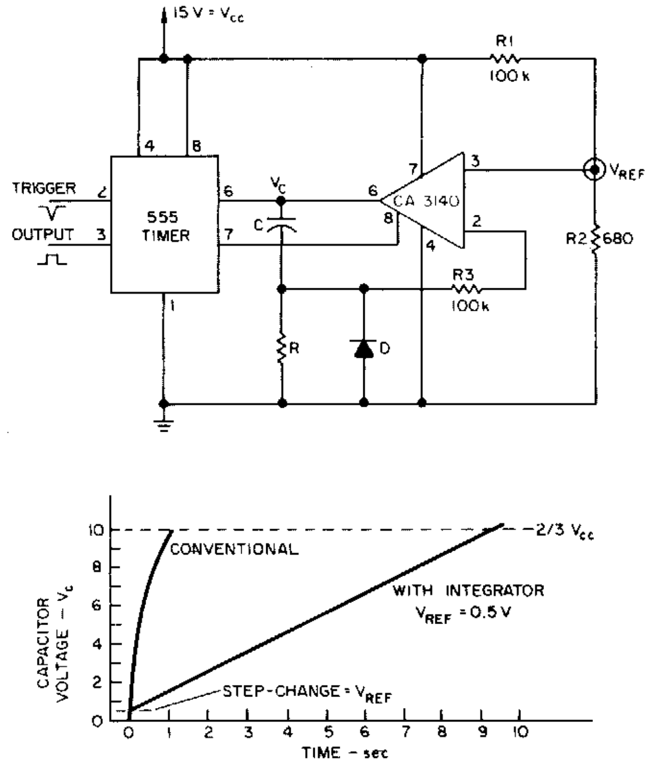


Fig. 105-2

MOTOROLA INC.

LOW-COST INTEGRATOR MULTIPLIES 555 TIMER'S DELAY

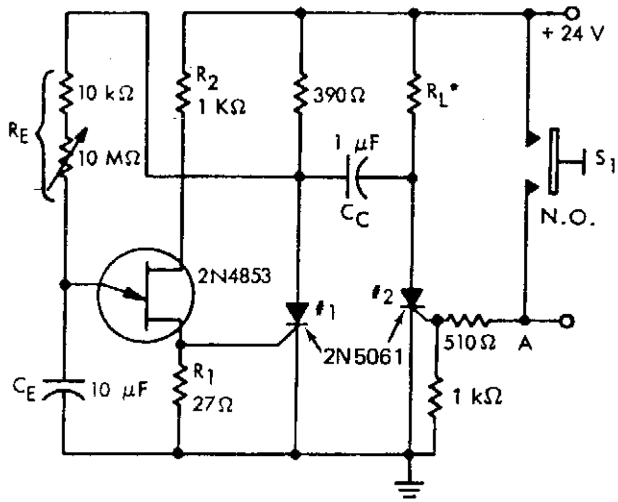


ELECTRONIC DESIGN ***Fig. 105-3***

Circuit Notes

Long delay times can be derived from a 555 timer with reasonably sized capacitors if an integrator circuit is used. The capacitor's charging time with an integrator circuit can be much longer than with a conventional 555-timer configuration.

SIMPLE TIME-DELAY CIRCUIT USING TWO SCRs



*Value of R_L must be low enough to allow hold current to flow in the SCR.

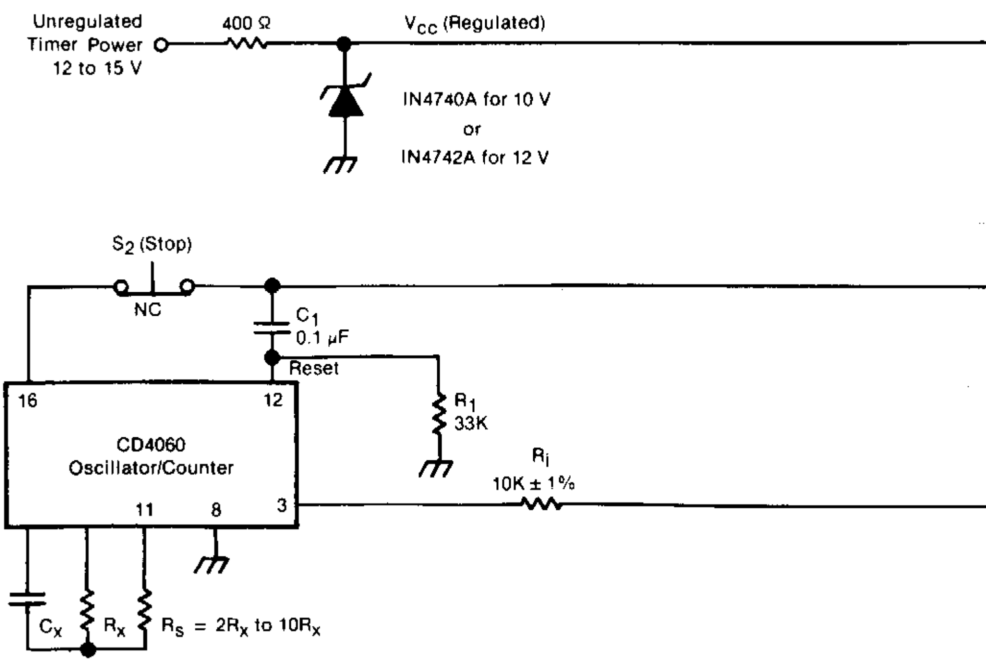
MOTOROLA INC. *Fig. 105-4*

106

Timers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LONG-TERM ELECTRONIC TIMER

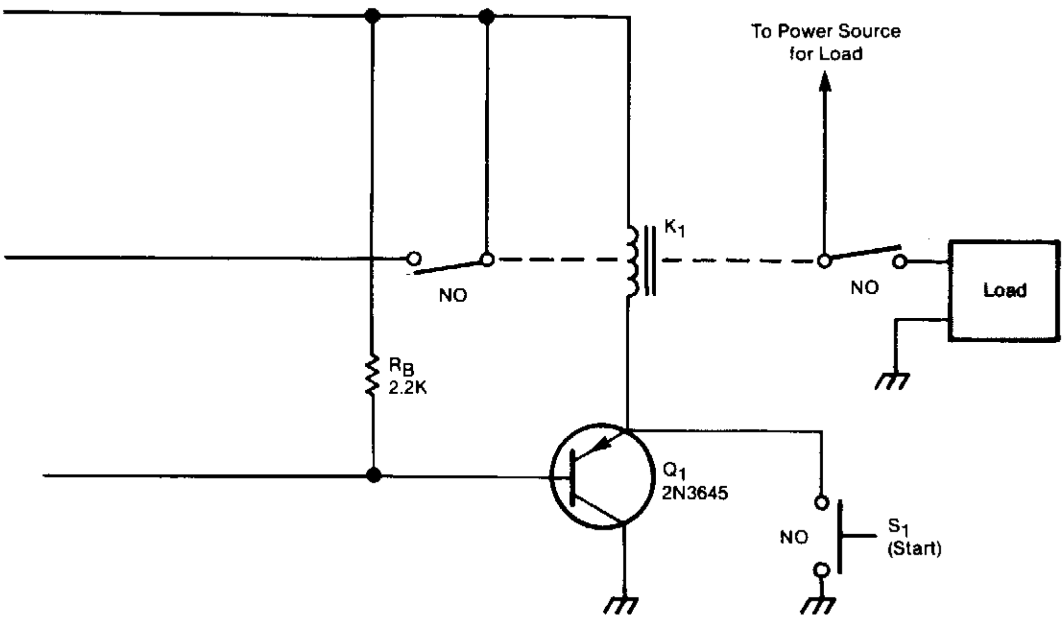


NASA

Fig. 106-1

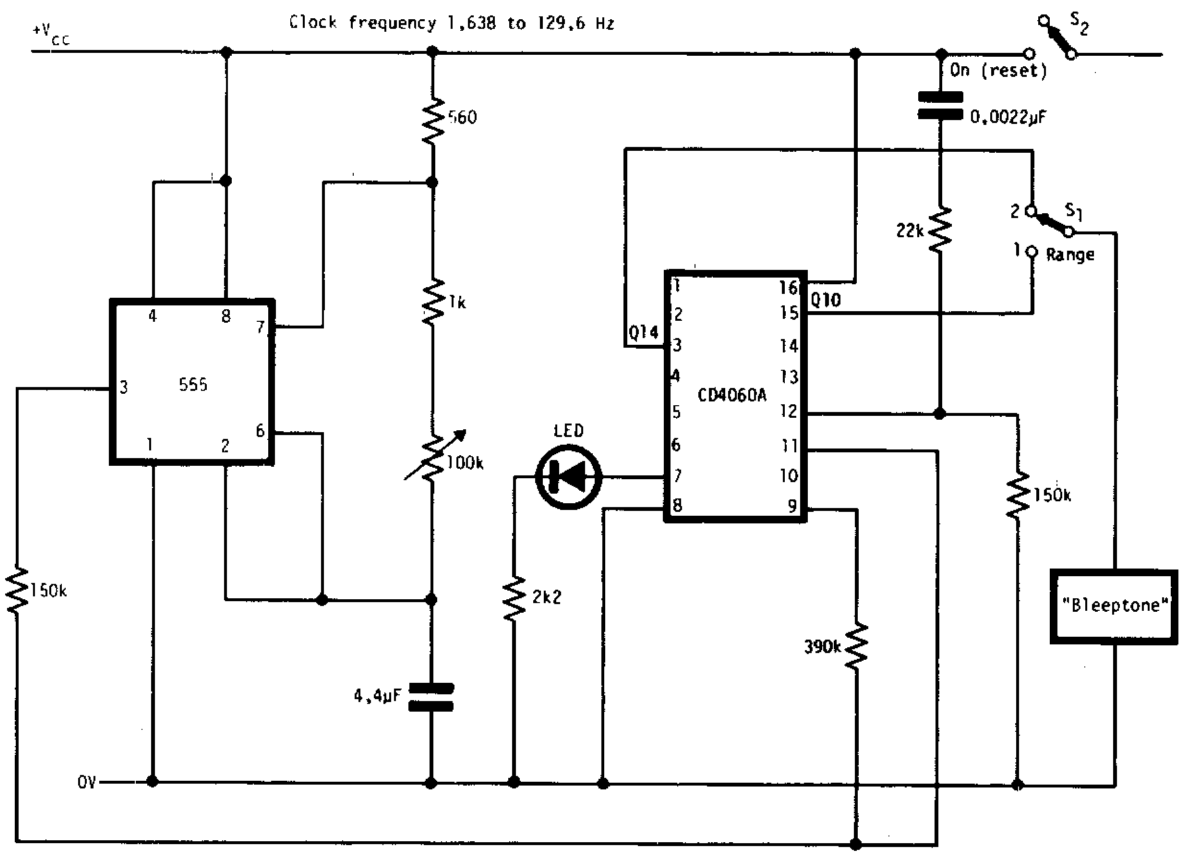
Circuit Notes

The timer includes an oscillator and a counter in an integrated circuit. The timing interval equals the oscillator period multiplied by the number of cycles to be counted. The oscillator frequency depends upon resistor RS and capacitor CX. The number of oscillator cycles to be counted before the counter output changes state is determined by the selection of the counter output terminal, shown here as pin 3. The interval can be set anywhere in the range from fractions of a second to months; it is given by $T = 0.55\ R_S C_X 2^n$, where n is an integer determined by the counter-output selection. Operation is initiated by the closure of momentary switch S1 (or by a command signal having a similar effect). This grounds one side of relay K1, thereby activating the relay



and causing the closure of the switches that supply power to the timer and to the load. The turn-on of V_{CC} at the timer is coupled through C1 to the counter-reset terminal, thus resetting the counter. The initial reset voltage transient is then drained away through R1 to permit normal operation. During the first half cycle of the counter operation, the counter output voltage (at pin 3 in this case) is low. This turns on transistor Q1 so that relay K1 latches on, enabling the timer to continue running even though switch S1 has opened. The oscillator runs while the relay is on. When the number of oscillator cycles reaches the limit, the counter output voltage at pin 3 goes high. This turns off Q1, thereby turning off the relay and returning the system to the original "power-off" state to await the next starting command. The timing cycle can also be interrupted and the system turned off by opening normally-closed switch S2.

TIMER WITH ALARM



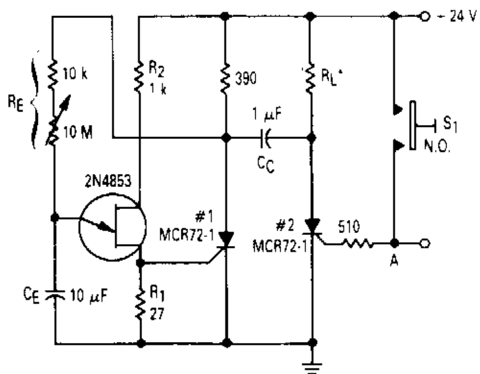
ELECTRONIC ENGINEERING

Fig. 106-2

Circuit Notes

The circuit has two ranges: 10 secs to 5 mins and 1 min to 80 mins. It can be powered by a 9-V battery. With the LED connected as shown a reasonable frequency of flashing occurs throughout the range of operation. This circuit is reset when S2 is closed.

TIMER CIRCUIT

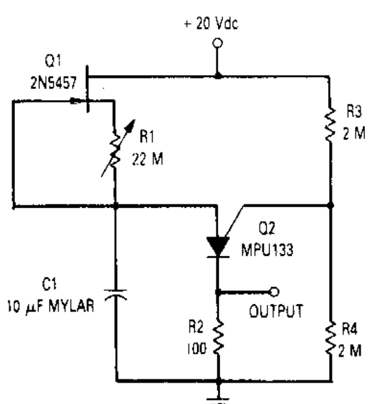


MOTOROLA *Fig. 106-3*

Circuit Notes

After one cycle of operation, SCR 1 will be on, and a low value of voltage is applied to the UJT emitter circuit, interrupting the timing function. When pushbutton S1 is pushed, or a positive going pulse is applied at point A, SCR 2 will turn on, and SCR 1 will be turned off by commutating capacitor CC. With SCR 1 off, the supply voltage will be applied to RE and the circuit will begin timing again. After a period of time determined by the setting of RE, the UJT will fire and turn SCR 1 on and commutate SCR 2 off. The time delay is determined by the charge time of the capacitor.

PUT LONG DURATION TIMER

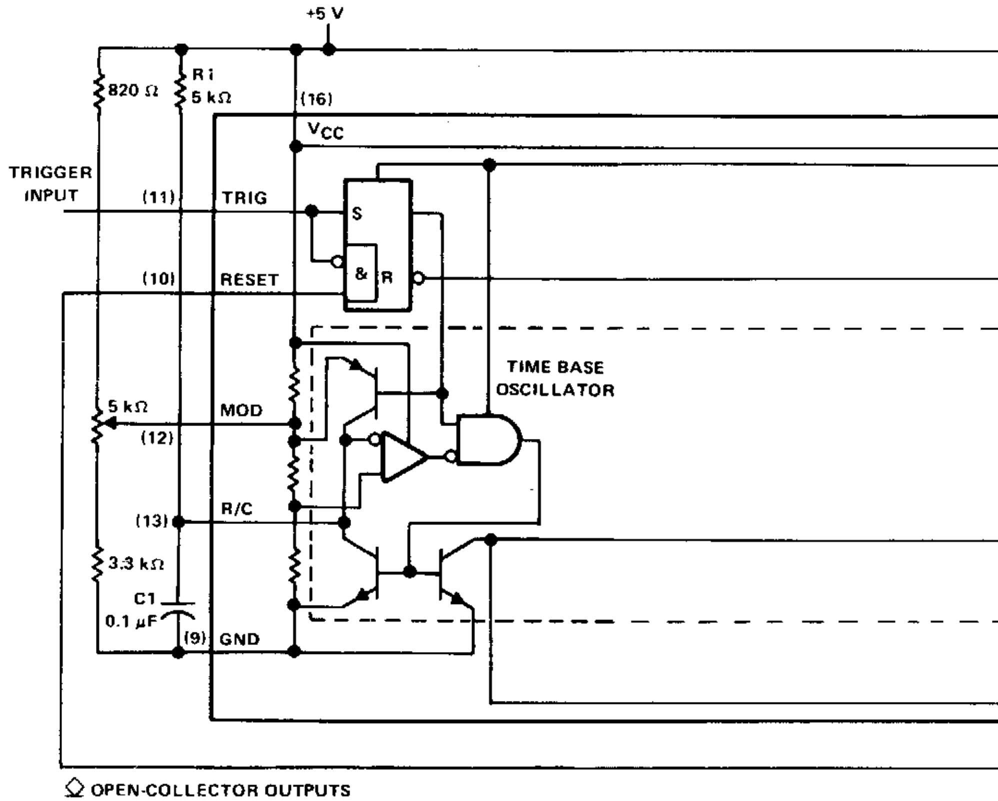


MOTOROLA *Fig. 106-4*

Circuit Notes

The time circuit can provide a time delay of up to 20 minutes. The circuit is a standard relaxation oscillator with a FET current source in which resistor R1 is used to provide reverse bias on the gate-to-source of the JFET. This turns the JFET off and increases the charging time of C1. C1 should be a low leakage capacitor such as a mylar type.

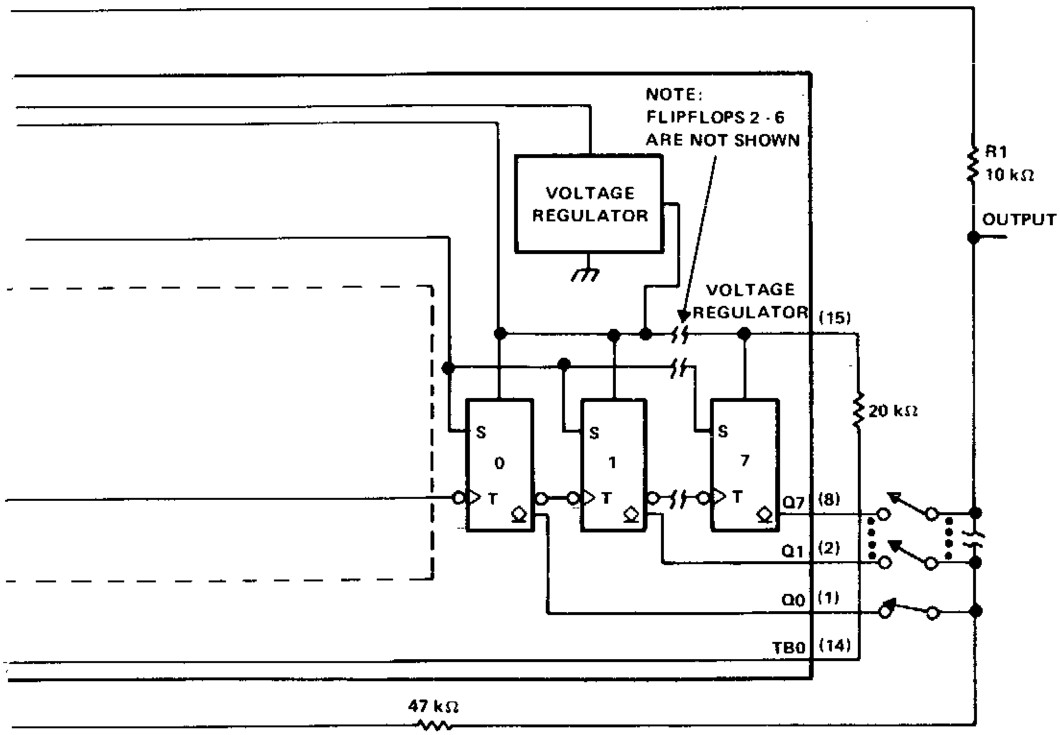
PROGRAMMABLE VOLTAGE CONTROLLED TIMER



TEXAS INSTRUMENTS *Fig. 106-5*

Circuit Notes

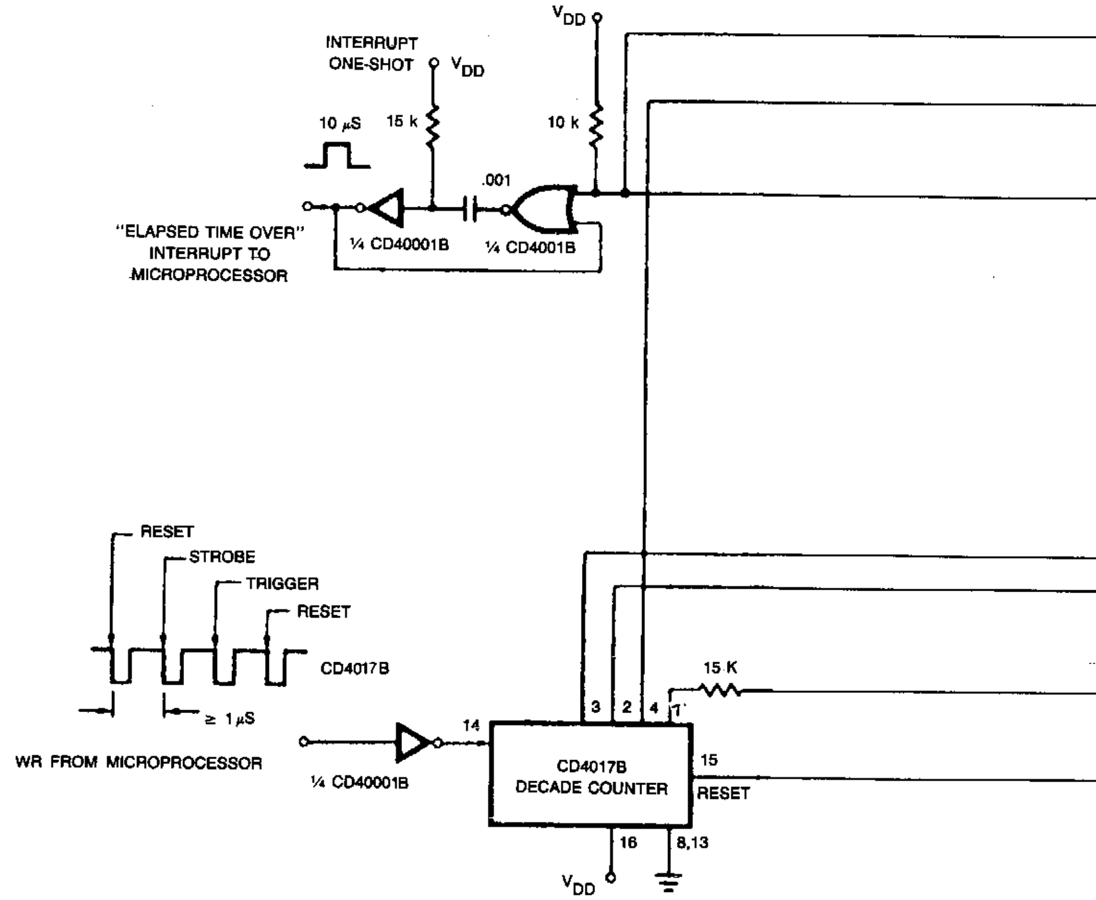
The μA2240 may easily be configured as a programmable voltage controlled timer with a minimum number of external components. The modulation input (pin 12), which allows external adjustment of the input threshold level. A variable voltage is applied from the arm of a 10 k ohm potentiometer connected from V_{CC} to ground. A change in the modulation input voltage will result in a change in the time base oscillator frequency and the period of the time base output (TBO). The TBO has an open-collector output that



is connected to the regulator output via a 10 k ohm pull-up resistor. The output of the TBO drives the input to the 8-stage counter section.

At start-up, a positive trigger pulse starts the TBO and sets all counter outputs to a low state. The binary outputs are open-collector stages that may be connected together to the 10 k ohm pull-up resistor to provide a "wired-OR" output function. This circuit may be used to generate 255 discrete time delays that are integer multiples of the time-base period. The total delay is the sum of the number of time-base periods, which is the binary sum of the Q outputs connected. Delays from 200 μs to 0.223 s are possible with this configuration.

LOW POWER MICROPROCESSOR PROGRAMMABLE INTERVAL TIMER

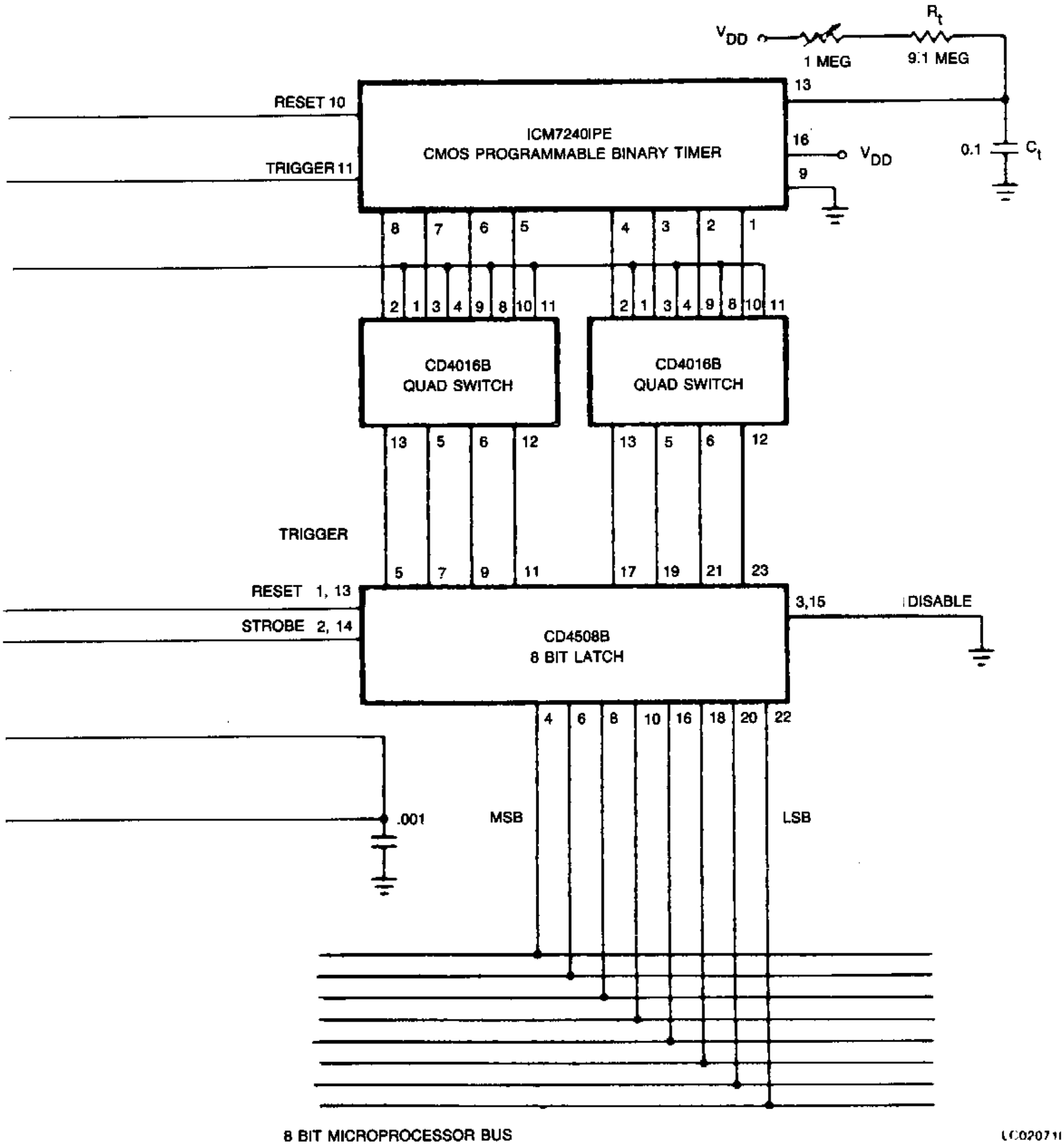


INTERSIL

Fig. 106-6

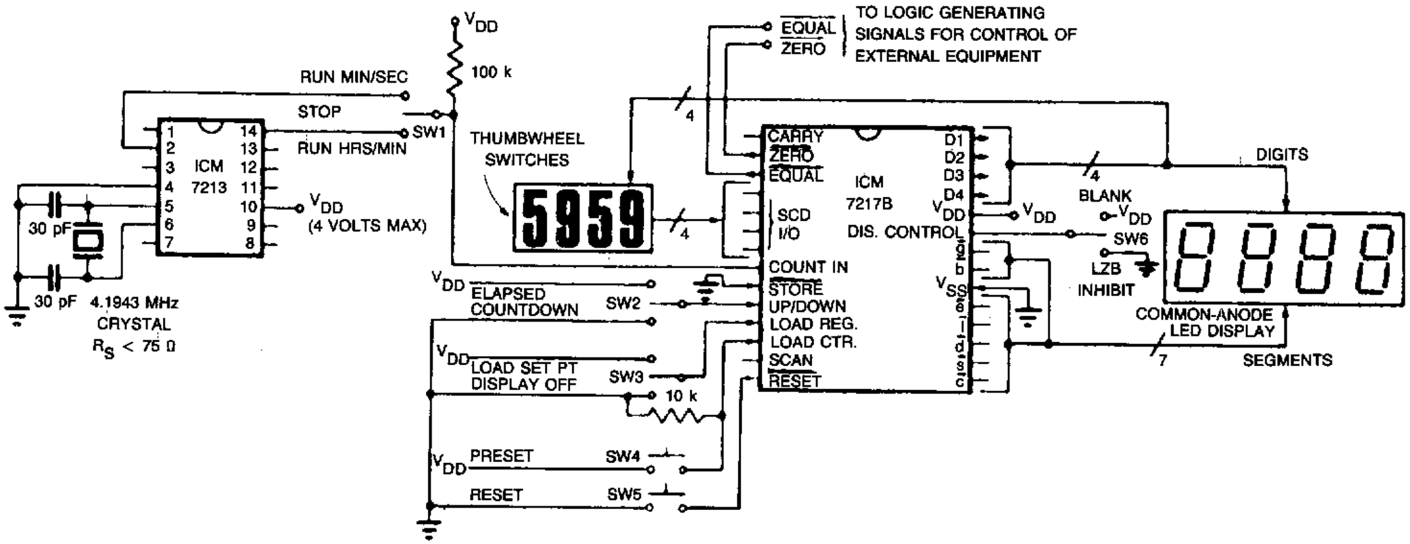
Circuit Notes

The microprocessor sends out an 8-bit binary code on its 8-bit I/O bus (the binary value needed to program the ICM7240), followed by four WRITE pulses into the CD4017B decade counter. The first pulse resets the 8-bit latch, the second strobes the binary value into the 8-bit latch, the third triggers the ICM7240 to begin its timing cycle and the fourth resets the decade counter. The ICM7240 then counts the interval of time



determined by the R-C value on pin 13, and the programmed binary count on pins 1 through 8. At the end of the programmed time interval, the interrupt one-shot is triggered, informing the microprocessor that the programmed time interval is over. With a resistor of approximately 10 M ohm and a can capacitor of 0.1 μF, the time base of the ICM7240 is one second. Thus, a time of 1-255 seconds can be programmed by the microprocessor, and by varying R or C, longer or shorter time bases can be selected.

PRECISION ELAPSED TIME/COUNTDOWN TIMER



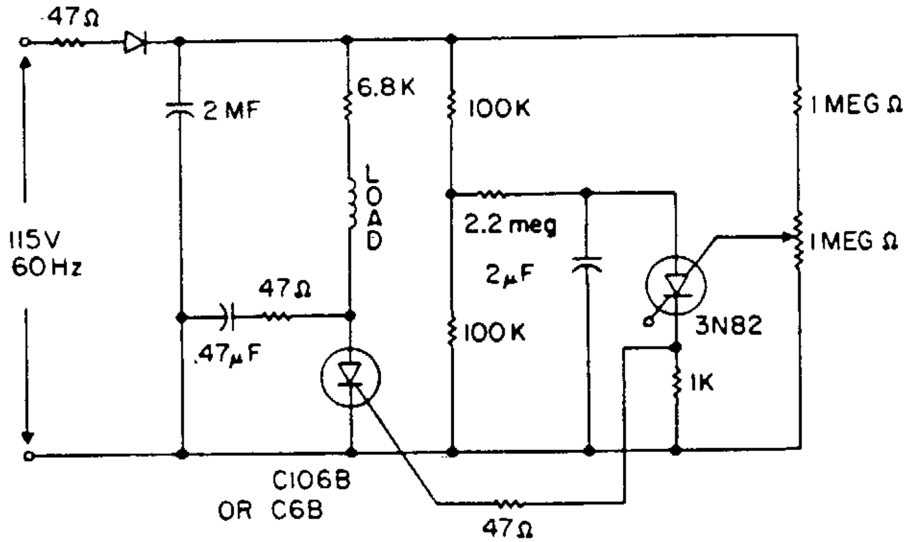
INTERSIL

Fig. 106-7

Circuit Notes

The circuit uses an ICM7213 precision one minute/one second timebase generator using a 4.1943 MHz crystal for generating pulses counted by an ICM7217B. The thumbwheel switches allow a starting time to be entered into the counter for a preset-countdown type timer, and allow the register to be set for compare functions. For instance, to make a 24-hour clock with BCD output the register can be preset with 2400 and the EQUAL output used to reset the counter. Note the 10 k resistor connected between the LOAD COUNTER terminal and ground. This resistor pulls the LOAD COUNTER input low when not loading, thereby inhibiting the BCD output drivers. This resistor should be eliminated and SW4 replaced with an SPDT center-off switch if the BCD outputs are to be used.

ADJUSTABLE AC TIMER .2 TO 10 SEC.



GENERAL ELECTRIC

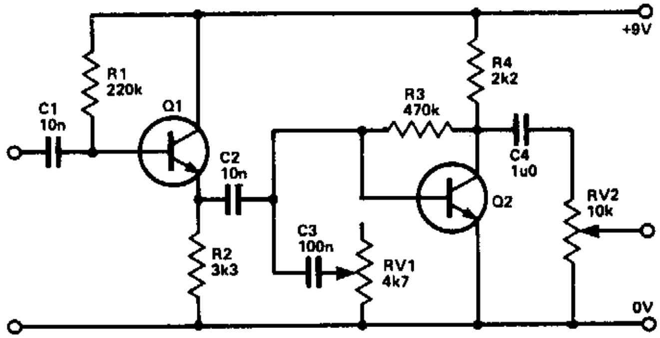
Fig. 106-8

107

Tone Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

GUITAR TREBLE BOOST

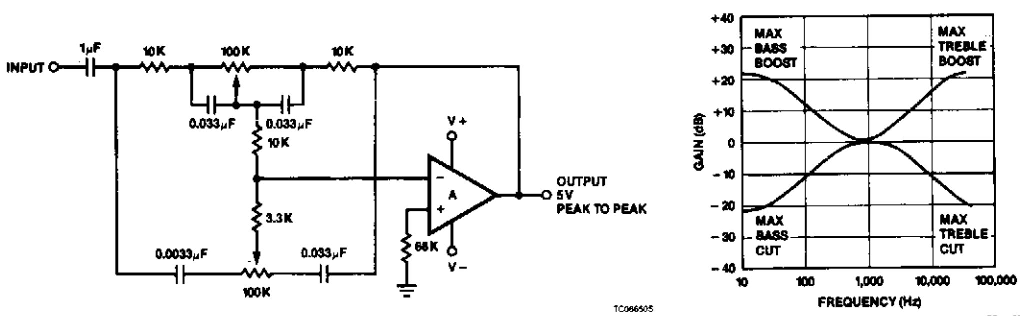


ELECTRONICS TODAY INTERNATIONAL *Fig. 107-1*

Circuit Notes

Q1 is connected as an emitter follower in order to present a high input impedance to the guitar. C2, being a relatively low capacitance, cuts out most of the bass, and C3 with RV1 acts as a simple tone control to cut the treble, and hence the amount of treble boost can be altered. Q2 is a simple preamp to recover signal losses in C2, C3, and RV1.

TONE CONTROL



NOTES:
1. Amplifier A may be a NE531 or 301. Frequency compensation, as for unity gain non-inverting amplifiers, must be used.
2. Turn-over frequency — 1kHz.
3. Bass boost + 20dB at 20Hz, bass cut − 20dB at 20Hz, treble boost + 19dB at 20Hz, treble cut − 19dB at 20Hz.

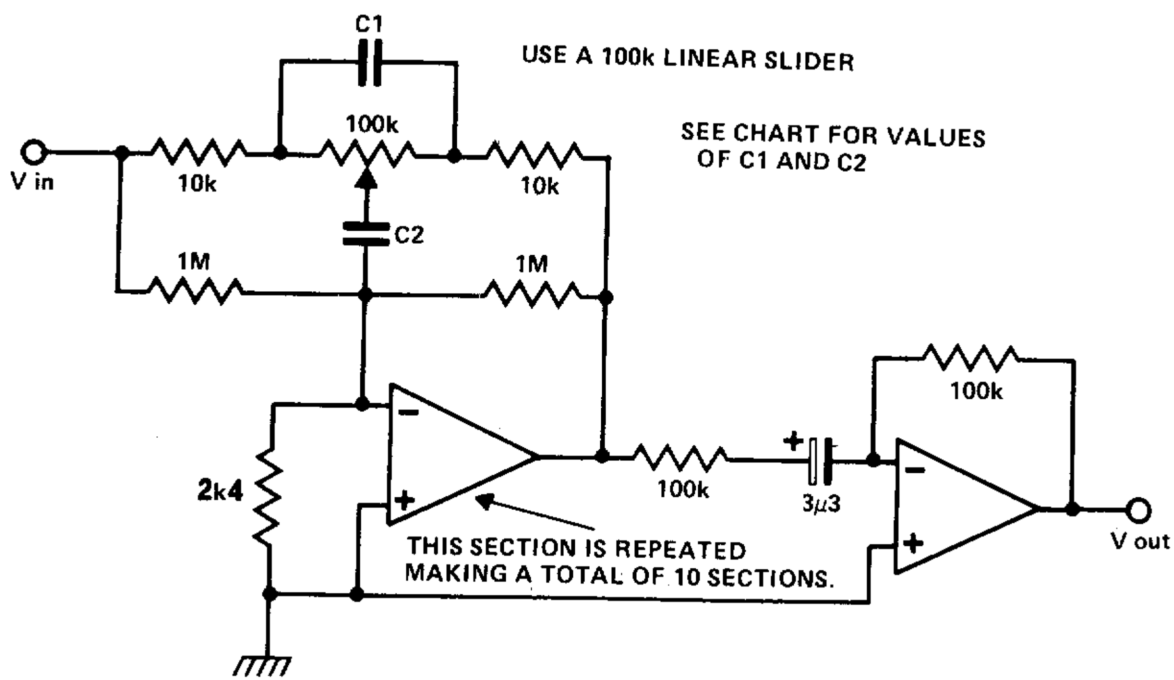
All resistor values are in ohms.

SIGNETICS *Fig. 107-2*

Circuit Notes

Tone control of audio systems involves altering the flat response in order to attain more low frequencies or more high ones, dependent upon listener preference. The circuit provides 20 dB of bass or treble boost or cut as set by the variable resistance. The actual response of the circuit is shown also.

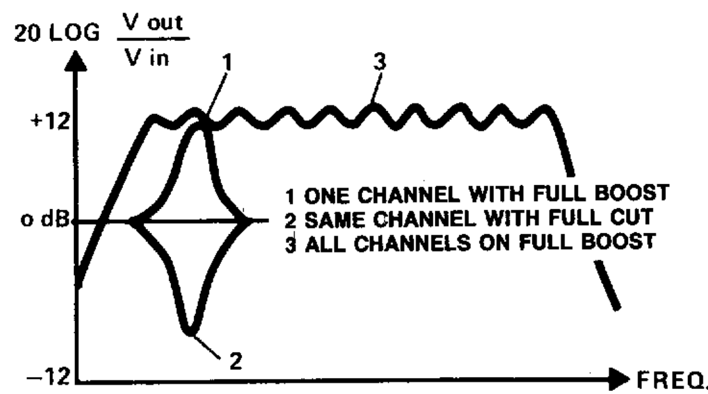
TEN BAND GRAPHIC EQUALIZER, USING ACTIVE FILTERS



ELECTRONICS TODAY INTERNATIONAL

Fig. 107-3

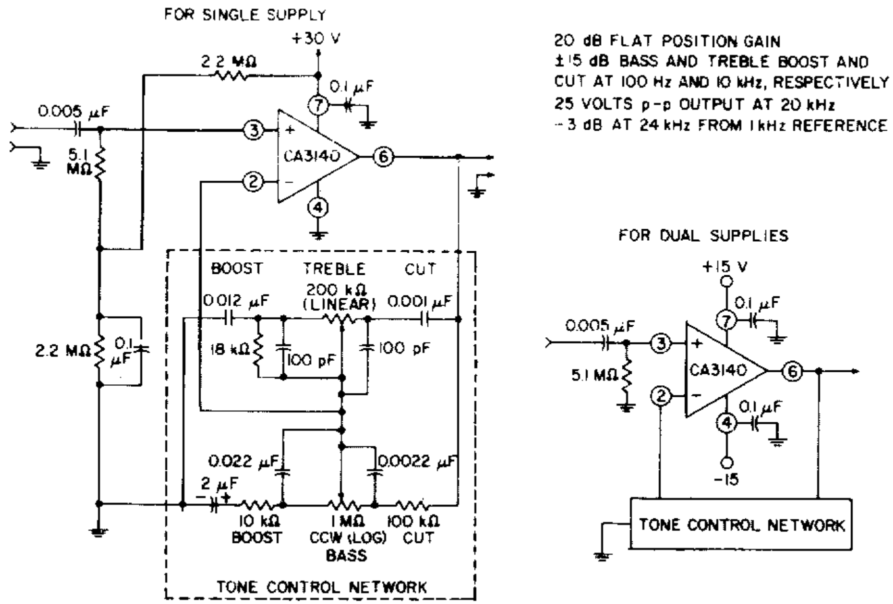
CHANNEL CENTRE FREQ. IN Hz	C1	C 2
32	180n	18n
64	100n	10n
125	47n	4n7
250	22n	2n2
500	12n	1n2
1000	5n6	560p
2000	2n7	270p
4000	1n5	150p
8000	680p	68p
16000	360p	36p



Circuit Notes

The above circuit is repeated ten times. Use the table to calculate values.

TONE-CONTROL AUDIO AMPLIFIER



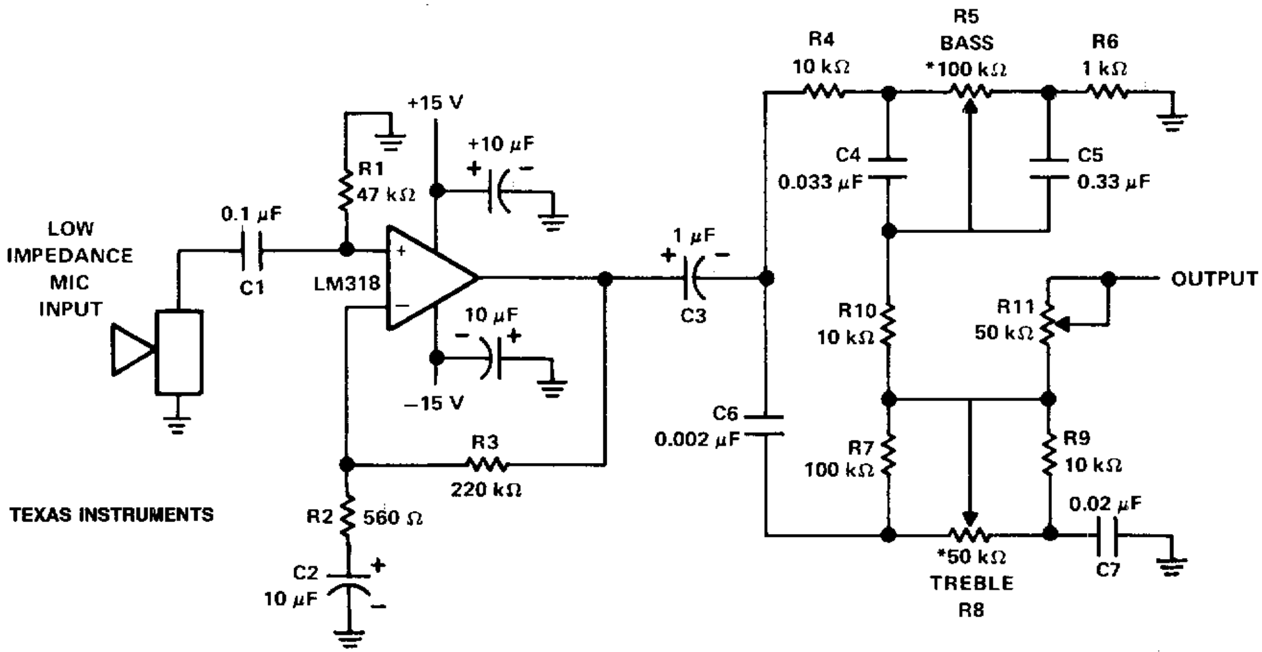
GENERAL ELECTRIC/RCA

Fig. 107-4

Circuit Notes

The circuit makes excellent use of the high slew rate, wide bandwidth, high input impedance, and high-output voltage capability of the CA3140 BiMOS op amp. The wideband gain of this circuit is equal to the ultimate boost or cut plus one, in this case a gain of eleven. For 20-dB boost or cut, input loading is essentially equal to the resistance from terminal 3 to ground.

MIKE PREAMP WITH TONE CONTROL



TEXAS INSTRUMENTS

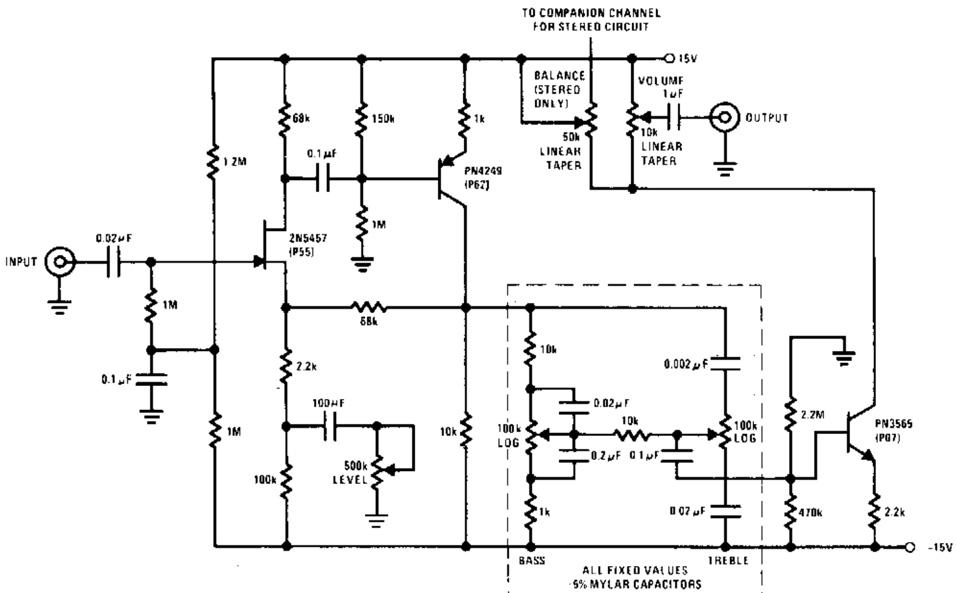
*THE TONE CONTROLS ARE AUDIO TAPER (LOG) POTENTIOMETERS.

Fig. 107-5

Circuit Notes

The LM318 op amp is operated as a standard non-inverting amplifier. Resistor R1 (47 k ohm) provides an input path to ground for the bias current of the non-inverting input. The combination of R2 (560 ohm) and C2 (10 μF) provides a frequency roll-off below 30 Hz. At 30 Hz and above the gain is relatively flat at about 50 dB, set by the ratio R3/R2. R3 (200 k ohm) furnishes negative feedback from the output to the inverting input of the op amp. C3 (1.0 μF electrolytic) ac couples the preamp to the tone control section. The top half of the tone control section is the bass control. The bottom half controls the treble frequency response. These tone controls (R5 and R8) require audio taper (logarithmic) potentiometers. The 50 k ohm potentiometer on the output can be used to set the output or gain of the preamp.

LOW COST HIGH-LEVEL PREAMP AND TONE CONTROL CIRCUIT



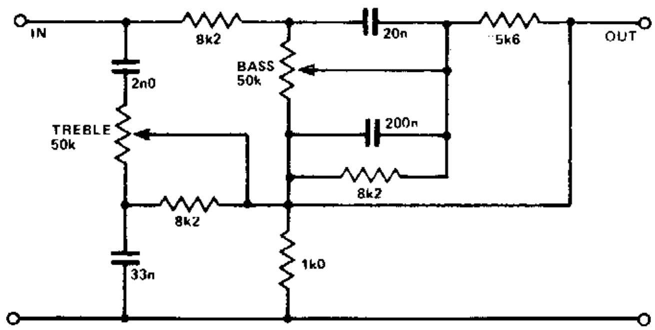
NATIONAL SEMICONDUCTOR CORP.

Fig. 107-6

Circuit Notes

This preamp and tone control uses the JFET to its best advantage; as a low noise high input impedance device. All device parameters are noncritical, yet the circuit achieves harmonic distortion levels of less than 0.05% with a S/N ratio of over 85 dB. The tone controls allow 18 dB of cut and boost; the amplifier has a 1-V output for 100-mV input at maximum level.

PASSIVE TONE-CONTROL CIRCUIT



ELECTRONICS TODAY INTERNATIONAL *Fig. 107-7*

Circuit Notes

A simple circuit using two potentiometers and easily available standard value components provides tone control. The impedance level is suitable for low-level transistor or op amp circuitry.

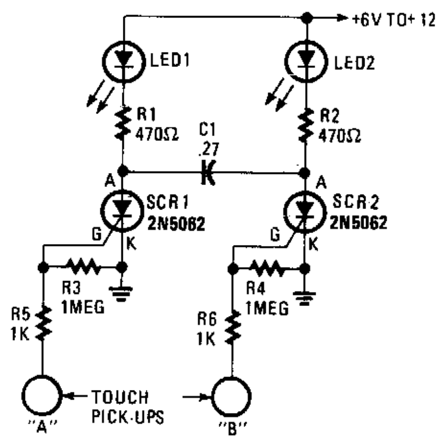
108

Touch-Switch Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Touch On/Off Switch
Touch Switch
Touchomatic

TOUCH ON/OFF SWITCH

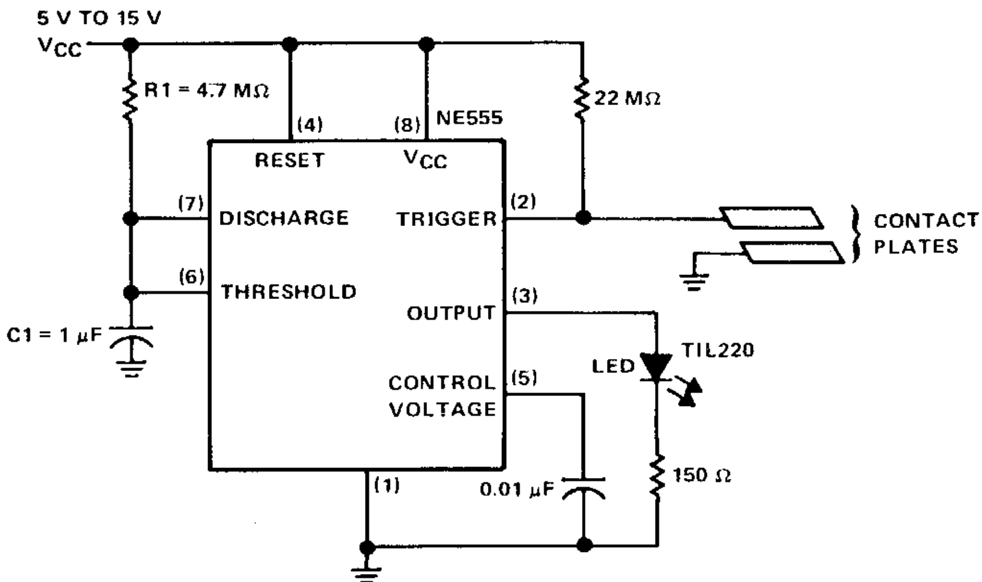


HANDS-ON ELECTRONICS *Fig. 108-1*

Circuit Notes

If a Touch On/Off Switch is desired, this circuit fills the bill. Two sensitive gate SCRs are interconnected, so that when one of the devices is turned on, the other (if on) is forced off. That toggling effect gives an on/off circuit condition for each of the LEDs in the SCR-anode circuits. To turn LED1 on and LED2 off, simply touch the "A" terminal, and to turn LED1 off and LED2 on, the "B" pick-up must be touched. It is possible to simultaneously touch both terminals, causing both SCRs to turn on together. To reset the circuit to the normal one-on/one-off condition, momentarily interrupt the circuit's dc power source. Additional circuitry can be connected to the anode circuit of either or both SCRs to be controlled by the on/off function of the touch switch.

TOUCH SWITCH



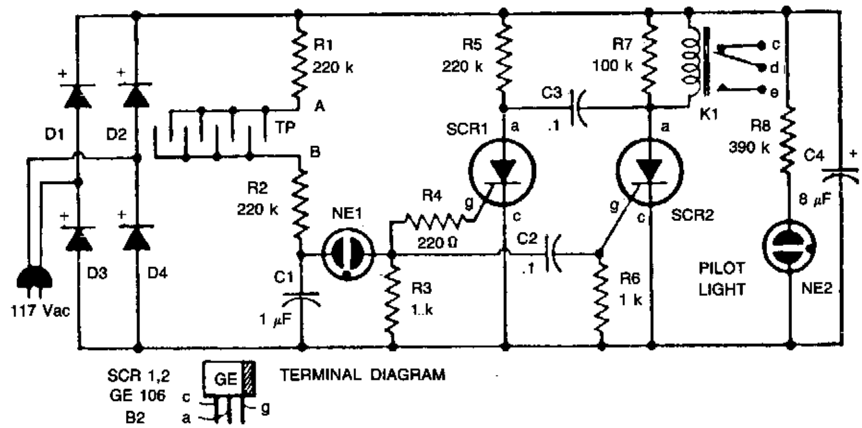
TEXAS INSTRUMENTS

Fig. 108-2

Circuit Notes

The circuit is basically a NE555 monostable, the only major difference being its method of triggering. The trigger input is biased to a high value by the 22 M ohm resistor. When the contact plates are touched, the skin resistance of the operator will lower the overall impedance from pin 2 to ground. This action will reduce the voltage at the trigger input to below the ⅓ V_{CC} trigger threshold and the timer will start. The output pulse width will be T = 1.1 R1C1, in this circuit about 5 seconds. A relay connected from pin 3 to ground instead of the LED and resistor could be used to perform a switching function.

TOUCHOMATIC



TAB BOOKS, INC.

Fig. 108-3

Circuit Notes

When someone touches the touchplate (TP), the resistance of his finger across points A and B is added in series to the combination of R1 and R2, the capacitor C2 begins to charge. When the voltage across C1 is finally sufficient to fire NE1, C1 will begin to discharge. When NE1 fires, it produces a short between its terminals. Since R3 is connected across C1, they are effectively in series after NE1 fires. A voltage spike will then be passed by C2 and this will act as a positive triggering pulse. The pulse is fed to both SCR gates: SCR2 conducts, thereby closing relay K1. With a finger no longer on the touchplate, no more pulses are forthcoming because the C1 charge path is open. The next contact with the touchplate will produce a pulse which triggers SCR1. SCR2 is now off by capacitor C3 which was charged by current passing through R6 and SCR2. The firing of SCR1 in this way places a negative voltage across SCR2 which momentarily drops the relay current to a point below the holding current value of SCR2. (Holding current is the minimum current an SCR requires to remain in a conducting state once its gate voltage is removed.) With SCR2 turned off, the relay will open and SCR1 will turn off due to the large resistance in series with its anode. Starved in this way SCR1 turns off because of a forced lack of holding current.

Sources Index

Chapter 1

Fig. 1-1. Signetics 1987 Linear Data Manual Vol. 2, 2/87, p.7-65.

Fig. 1-2. General Electric Application Note 90.16, p. 25.

Fig. 1-3. Courtesy, William Sheets.

Fig. 1-4. General Electric Application Note 90.16, p. 25.

Fig. 1-5. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 27.

Fig. 1-6. R-E Experimenters Handbook, p. 157.

Chapter 2

Fig. 2-1. Linear Technology Corporation, Linear Applications Handbook, 1987, p. AN21-2

Fig. 2-2. Linear Technology Corporation, Linear Databook Supplement, 1988, p. S2-34.

Fig. 2-3. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-154.

Fig. 2-4. Linear Technology Corp., Linear Databook, 1986, p. 2-83.

Fig. 2-5. National Semiconductor Corp., Transistor Databook, 1982, p. 11-23.

Fig. 2-6. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN21-1.

Fig. 2-7. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 10/86, p. 4-260.

Fig. 2-8. Electronic Engineering, 11/86, p. 40.

Fig. 2-9. National Semiconductor Corp., Transistor Databook, 1982, p. 11-25.

Fig. 2-10. Siliconix, Integrated Circuits Data Book, 3/85, p. 2-112.

Fig. 2-11. Electronics Engineering, 9/78, p. 17.

Fig. 2-12. Electronics Engineering, 9/84, p. 33.

Fig. 2-13. Signetics, 1987 Linear Data Manual, Vol. 1: Communications, 8/87, p. 4-346.

Fig. 2-14. NASA Tech Briefs, Spring 1983, p. 244.

Fig. 2-15. Radio Electronics, 7/83, p. 74.

Fig. 2-16. Motorola Inc., Linear Integrated Circuits, 1979, p. 6-58.

Fig. 2-17. MR-E Experimenters Handbook, p. 158.

Fig. 2-19. Popular Electronics, 8/68.
Fig. 2-20. Ham Radio, 9/84, p. 24.
Fig. 2-21. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 20.
Fig. 2-22. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-2, 3-4.
Fig. 2-23. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 3/87, p. 4-345.
Fig. 2-24. Electronic Engineering, 11/85, p. 32.
Fig. 2-25. NASA Tech Briefs, Sept/Oct 1986, p. 43.

Chapter 3

Fig. 3-1. Intersil, Component Data Catalog, 1987, p. 4-43.
Fig. 3-2. Linear Technology, Application Note 9, p. 16.
Fig. 3-3. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 26.
Fig. 3-4. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN15-2.
Fig. 3-5. Linear Technology Corp., Linear Databook, 1986, p. 5-17.
Fig. 3-6. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 11/86, p. 5-215.
Fig. 3-7. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S5-126.
Fig. 3-8. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 12/86, p. 4-67.

Chapter 4

Fig. 4-1. Electronic Design, 5/82, p. 214.
Fig. 4-2. R-E Experimenters Handbook, p. 160.
Fig. 4-3. Popular Electronics, 11/73, p. 50.

Chapter 5

Fig. 5-1. Hands-On Electronics, Summer 1984, p. 77.
Fig. 5-2. Courtesy, William Sheets.
Fig. 5-3. Courtesy, William Sheets.
Fig. 5-4. Electronic Engineering, 5/84, p. 44
Fig. 5-5. Texas Instruments, Linear and Interface Circuits Applications, vol. 1, 1985, p. 3-13.
Fig. 5-6. Electronics Today International, 4/85, p. 82.

Chapter 6

Fig. 6-1. Tab Books Inc., 101 Sound, Light, and Power IC Projects.
Fig. 6-2. Radio Electronics, 7/70, p. 38.
Fig. 6-3. Hands-On Electronics, Jul/Aug 1986, p. 16.
Fig. 6-4. 73 Magazine, 12/76, p. 170.
Fig. 6-5. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 21.
Fig. 6-6. Ibid.
Fig. 6-7. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-17.
Fig. 6-8. Signetics, Analog Data Manual, 1982, p. 3-90.

Chapter 7

Fig. 7-1. Motorola, TMOS Power FET Design Ideas, 1985.
Fig. 7-2. Hands-On Electronics, 4/87, p. 95.
Fig. 7-3. Tab Books, Inc., The Giant book of Easy-To-Build Electronics Projects, 1982, p. 196.
Fig. 7-4. General Electric, Optoelectronics, Third Edition, p. 151.
Fig. 7-5. Tab Books, Inc., The Build-It Book of Electronic Projects, No. 1498, p. 28.
Fig. 7-6. Radio-Electronics, 6/85, p. 60.
Fig. 7-7. Popular Electronics, 12/74, p. 6.
Fig. 7-8. Popular Electronics, 4/75, p. 68.
Fig. 7-9. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 10-21.
Fig. 7-10. Texas Instruments, Linear and Interface Circuits Applications, 1985, vol. 1, p. 3-5.
Fig. 7-11. Radio-Electronics, 1979.
Fig. 7-12. Hands-On Electronics, 1/87, p. 30.
Fig. 7-13. Electronic Engineering, 12/75, p. 9.
Fig. 7-14. Electronic Design 18, 9/76, p. 114.
Fig. 7-15. Radio-Electronics, 5/87, p. 10.
Fig. 7-16. Hands-On Electronics, Fall 1984, p. 45.
Fig. 7-17. Hands-On Electronics, 4/87, p. 92.
Fig. 7-18. General Electric, Optoelectronics, Third Edition, p. 105.

Chapter 8

Fig. 8-1. CQ, 7/82, p. 18.
Fig. 8-2. Electronics Design, 7/76, p. 120.
Fig. 8-3. Courtesy, William Sheets.
Fig. 8-4. Moli Energy Limited.
Fig. 8-5. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN6-3.
Fig. 8-6. Courtesy, William Sheets.
Fig. 8-7. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, p. 6-24.
Fig. 8-8. Linear Technology Corp., Linear Databook Supplement, 1988, p. S5-11.
Fig. 8-9. Motorola, TMOS Power FET Design Ideas, 1985, p. 8.
Fig. 8-10. Radio Electronics, 9/85, p. 44.
Fig. 8-11. Siliconix, MOSpower Applications Handbook, p. 6-176.

Chapter 9

Fig. 9-1. Motorola, TMOS Power FET Ideas, 1985, p. 7.
Fig. 9-2. Electronic Engineering, 2/85, p. 45.
Fig. 9-3. Electronic Engineering, 10/70, p. 17.
Fig. 9-4. Electronic Engineering, Mid5/78, p. 11.
Fig. 9-5. Moli Energy Limited, Publication MEL-126.
Fig. 9-6. Linear Technology Corp., Linear Databook, 1986, p. 2-104.
Fig. 9-7. Courtesy, William Sheets.
Fig. 9-8. Tab Books, Inc., 101 Sound, Light, and Power IC Projects.

Chapter 10

Fig. 10-1. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 2/87, p. 5-367.
Fig. 10-2. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-3, 3-4.
Fig. 10-3. Courtesy, William Sheets.
Fig. 10-4. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 17.
Fig. 10-5. Siliconix, Small-Signal FET Data Book, 1/86, p. 7-29.
Fig. 10-6. Linear Technology Corp., Linear Databook, 1986, p. 2-83.
Fig. 10-7. Linear Technology Corp., Linear Databook, 1986, p. 2-101.

Chapter 11

Fig. 11-1. Signetics, 1987 Linear Data Manual, Vol. 1: Communications, 2/87, p. 4-312.
Fig. 11-2. Electronic Engineering, 5/86, p. 50.
Fig. 11-3. Electronic Design, 9/73, p. 148.
Fig. 11-4. Intersil, Component Data Catalog, 1987, p. 6-28.
Fig. 11-5. Signetics, 555 Timers, 1973, p. 19.

Chapter 12

Fig. 12-1. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 7-21.
Fig. 12-2. Electronic Engineering, 2/85, p. 34.

Chapter 13

Fig. 13-1. Hands-On Electronics, 3/87, p. 25.
Fig. 13-2. Motorola, TMOS Power FET Design Ideas, 1985, p. 17.
Fig. 13-3. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN13-23.
Fig. 13-4. Intersil, Component Data Catalog, 1987, p. 5-113.
Fig. 13-5. Courtesy, William Sheets.
Fig. 13-6. Linear Technology Corp., Linear Databook, 1986, p. 3-23.
Fig. 13-7. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-57.

Chapter 14

Fig. 14-1. Electronic Engineering, 3/78, p. 38.
Fig. 14-2. Hands-On Electronics, Nov/Dec 1985, p. 4.

Chapter 15

Fig. 15-1. Linear Technology Corp., Linear Databook, 1986, p. 2-82.
Fig. 15-2. Popular Electronics, 9/77, p. 92.
Fig. 15-3. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 5-269.
Fig. 15-4. Siliconix, Integrated Circuits Data Book, 3/85, p. 5-8.
Fig. 15-5. Electronic Engineering, 2/85, p. 45.
Fig. 15-6. Electronic Engineering, 2/84, p. 36.
Fig. 15-7. Electronic Design 15, 7/79, p. 120.
Fig. 15-8. Electronic Engineering, 12/78, p. 17.

Fig. 15-9. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 23.
Fig. 15-10. Electronic Engineering, 11/86, p. 39.
Fig. 15-11. Popular Electronics, 3/79, p. 77.
Fig. 15-12. Electronic Engineering, 7/86, p. 27.
Fig. 15-13. Electronic Engineering, 1/86, p. 37.

Chapter 16

Fig. 16-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 5-215.
Fig. 16-2. Linear Technology Corp., Linear Databook, 1986, p. 4-15.
Fig. 16-3. Siliconix, Integrated Circuits Data Book, 3/85, p. 2-207.
Fig. 16-4. Siliconix, Integrated Circuits Data Book, 3/85, p. 2-231.
Fig. 16-5. Intersil, Component Data Catalog, 1987, p. 13-51.
Fig. 16-6. Siliconix, Integrated Circuits Data Book, 3/85, p. 3-62.
Fig. 16-7. Datel, Data Conversion Components, p. 4-37.

Chapter 17

Fig. 17-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-62.
Fig. 17-2. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 4-136.
Fig. 17-3. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 5-269.
Fig. 17-4. Electronic Engineering, 12/77, p. 19.
Fig. 17-5. Siliconix, Integrated Circuits Data Book, 3/85, p. 5-17.
Fig. 17-6. Linear Technology Corp., Linear Databook, 1986, p. 5-17.
Fig. 17-7. Analog Devices, Data Acquisition Databook, 1982, p. 4-56.
Fig. 17-8. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 4-311.
Fig. 17-9. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 20.
Fig. 17-10. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 11.
Fig. 17-11. Siliconix, MOSpower Applications Handbook, p. 6-178.
Fig. 17-12. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 5-368.
Fig. 17-13. National Semiconductor Corp., Transistor Databook, 1982, p. 7-27.
Fig. 17-14. National Semiconductor Crop., 1984 Linear Supplemental Databook, p. S5-142.
Fig. 17-15. Electronic Engineering, 8/83, p. 141.

Chapter 18

Fig. 18-1. Intersil, Component Data catalog, 1987, p. 14-70.
Fig. 18-2. General Electric, Application Note 90.16, p. 29.
Fig. 18-3. General Electric, Application Note 90.16, p. 29.
Fig. 18-4. General Electric, Application Note 90.16, p. 28.
Fig. 18-5. Intersil, Component Data Catalog, 1987, p. 14-91.
Fig. 18-6. Intersil, Component Data Catalog, 1987, p. 7-96.
Fig. 18-7. Intersil, Component Data Catalog, 1987, p. 14-121.
Fig. 18-8. Intersil, Databook 1987, p. 7-47.

Chapter 19

Fig. 19-1. Ham Radio, 6/85, p. 23.
Fig. 19-2. Siliconix, MOSpower Design Catalog, 1/83, p. 5-27.
Fig. 19-3. Linear Technology Corp., Linear Databook, 1986, p. 2-104.
Fig. 19-4. QST, 12/85, p. 38.
Fig. 19-5. Linear Technology Corp., Linear Applications Handbook 1987, p. AN20-12.
Fig. 19-6. QST, 2/28, p. 43.
Fig. 19-7. Ham Radio, 2/79, p. 40.
Fig. 19-8. Courtesy, William Sheets.
Fig. 19-9. Courtesy, William Sheets.
Fig. 19-10. Electronic Design 21, 10/75, p. 98.
Fig. 19-11. Ham Radio, 2/79, p. 40.
Fig. 19-12. Ham Radio, 2/79, p. 42.
Fig. 19-13. Tab Books, Inc., The Complete Handbook of Amplifiers, Oscillators, and Multivibrators, No. 1230, p. 328.
Fig. 19-14. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 5-269.

Fig. 19-15. Motorola, MECL System Design Handbook, 1983, p. 227.
Fig. 19-16. Electronic Design, 11/69, p. 109.
Fig. 19-17. QST, 1/86, p. 40.
Fig. 19-18. R-E Experimenters Handbook, p. 157.
Fig. 19-19. Electronic Design 23, 11/74, p. 148.

Chapter 20

Fig. 20-1. Linear Technology Corp., 1986 Linear Databook, p. 2-57.
Fig. 20-2. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-13.
Fig. 20-3. Intersil, Component Data Catalog, 1987, p. 7-4.
Fig. 20-4. NASA Tech Briefs, Jul/Aug 1986, p. 37.
Fig. 20-5. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 17.
Fig. 20-6. Linear Technology Corp., Linear Databook, 1986, p. 2-85.
Fig. 20-7. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 14.

Chapter 21

Fig. 21-1. EXAR, Telecommunications Databook, 1986, p. 9-23.
Fig. 21-2. National Semiconductor Corp., Audio/Radio Handbook, 1980, p. 3-17.
Fig. 21-3. Signetics, 1987 Linear Data Manual, Vol. 1: Communications, 2/87, p. 4-66.
Fig. 21-4. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/86, p. 4-263.

Chapter 22

Fig. 22-1. Radio-Electronics, 12/86, p. 57.
Fig. 22-2. Radio-Electronics, 8/87, p. 63.
Fig. 22-3. Radio-Electronics, 8/87, p. 53.
Fig. 22-4. Radio-Electronics, 6/87, p. 12.
Fig. 22-5. EXAR, Telecommunications Databook, 1986, p. 9-23.
Fig. 22-6. Radio-Electronics, 3/86, p. 51.
Fig. 22-7. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/86, p. 7-123.
Fig. 22-8. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/86, p. 7-123.
Fig. 22-9. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/86, p. 4-295.

Chapter 23

Fig. 23-1. General Electric, Application Note 90.16, p. 26.
Fig. 23-2. Siliconix, Integrated Circuits Data Book, 3/85, p. 5-16.
Fig. 23-3. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-23.
Fig. 23-4. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 18.
Fig. 23-5. Intersil, Component Data Catalog, 1987, p. 7-44.
Fig. 23-6. Electronic Engineering, 11/86, p. 39.
Fig. 23-7. General Electric, Application Note 90.16, p. 27.
Fig. 23-8. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 5-367.
Fig. 23-9. Intersil, Component Data Catalog, 1987, p. 5-112.
Fig. 23-10. GENERAL Electric, Application Note 90.16, p. 26.

Chapter 24

Fig. 24-1. Electronic Engineering, 8/85, p. 30.
Fig. 24-2. Electronic Engineering, 11/86, p. 40.

Chapter 25

Fig. 25-1. CQ, 1/87, p. 36.

Chapter 26

Fig. 26-1. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 26.
Fig. 26-2. National Semiconductor Corp., Linear Databook, 1982, p. 171.
Fig. 26-3. Electronic Engineering, 9/84, p. 30.
Fig. 26-4. GENERAL Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 12.

Chapter 27

Fig. 27-1. Motorola, TMOS Power FET Design Ideas, 1985, p. 18.
Fig. 27-2. Electronic Design, 12/87, p. 67.
Fig. 27-3. General Electric/RCA, BIMOS Operational Amplifiers Circuit Ideas, 1987, p. 22.
Fig. 27-4. Electronic Engineering, 6/78, p. 32.
Fig. 27-5. Electronic Engineering, 2/83, p. 37.

Chapter 28

Fig. 28-1. Tab Books, Inc., The Giant Book of Easy-To-Build Electronic Projects, 1982, p. 53.
Fig. 28-2. Hands-On Electronics, 2/87, p. 38.

Chapter 29

Fig. 29-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 4-135.
Fig. 29-2. Linear Technology Corp., Linear Databook, 1986, p. 2-82.
Fig. 29-3. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/86, p. 4-135.
Fig. 29-4. Transistor Databook, 1982, p. 11-25.

Chapter 30

Fig. 30-1. Hands-On Electronics, May/Jun 1986, p. 52.
Fig. 30-2. Popular Electronics, 3/67.
Fig. 30-3. Courtesy, William Sheets.

Chapter 31

Fig. 31-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN13-22.
Fig. 31-2. General Electric, Optoelectronics, Third Edition, p. 149.
Fig. 31-3. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-39.

Chapter 32

Fig. 32-1. Courtesy, William Sheets.
Fig. 32-2. Courtesy, William Sheets.
Fig. 32-3. Hands-On Electronics, 8/87, p. 65.
Fig. 32-4. Courtesy, William Sheets.
Fig. 32-5. Hands-On Electronics, 3/87, p. 27.
Fig. 32-6. Ham Radio, 9/86, p. 67.

Chapter 33

Fig. 33-1. Electronic Engineering, 10/48, p. 45.
Fig. 33-2. Electronics Today International, 10/78, p. 26.
Fig. 33-3. Hybrid Products Databook, 1982, p. 17-131.
Fig. 33-4. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-62.
Fig. 33-5. Intersil, Component Data Catalog, 1987, p. 8-102.
Fig. 33-6. 73 for Radio Amateurs, 2/86, p. 10.
Fig. 33-7. Intersil, Component Data Catalog, 1987, p. 7-45.
Fig. 33-8. Linear Technology Corp., 1986 Linear Databook, p. 2-56.
Fig. 33-9. Electronic Engineering, 2/47, p. 47.
Fig. 33-10. Courtesy, William Sheets.
Fig. 33-11. Raytheon, Linear and Integrated Circuits, 1984, p. 6-205.
Fig. 33-12. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-7.
Fig. 33-13. Motorola, Linear Integrated Circuits, 1979, p. 3-147.
Fig. 33-14. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-9.

Chapter 34

Fig. 34-1. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-52.
Fig. 34-2. Popular Electronics, 3/81, p. 100.
Fig. 34-3. Hands-On Electronics, Spring 1986, p. 4.
Fig. 34-4. Hands-On Electronics, Fall 1984, p. 61.
Fig. 34-5. General Electric, SCR Manual, Fourth Edition, p. 85.
Fig. 34-6. Electronic Design 65, 3/73, p. 84.
Fig. 34-7. Electronic Design, 3/69, p. 96.
Fig. 34-8. Electronic Engineering, 676, p. 32.
Fig. 34-9. Popular Electronics, 3/75, p. 78.
Fig. 34-10. General Electric, Optoelectronics, Third Edition.
Fig. 34-11. Radio-Electronics, 2/87, p. 36.
Fig. 34-12. General Electric, Application Note 200.35, p. 16.
Fig. 34-13. General Electric, Application Note 90. 16, p. 27.

Fig. 34-14. National Semiconductor Corp., CMOS Databook, 1981, p. 8-45.
Fig. 34-15. General Electric, Application Note 90.25.
Fig. 34-16. Motorola, Circuit Applications for the Trian (AN-466), p. 11.
Fig. 34-17. Siliconix, MOSpower Applications Handbook, p. 6-181.
Fig. 34-18. Popular Electronics, 3/75, p. 78.

Chapter 35

Fig. 35-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN5-6.
Fig. 35-2. Linear Technology Corp., Linear Databook, 1986, p. 2-82.

Chapter 36

Fig. 36-1. Courtesy, William Sheets.
Fig. 36-2. Electronics Today International, 6/76, p. 43.
Fig. 36-3. Radio-Electronics, 4/87, p. 48.
Fig. 36-4. Radio-Electronics, 2/84, p. 97.
Fig. 36-5. Hands-On Electronics, Sep/Oct 1986, p. 24.
Fig. 36-6. Electronic Engineering, 9/86, p. 37.
Fig. 36-7. R-E Experimenters Handbook, p. 162.
Fig. 36-8. Linear Technology Corp., Linear Databook, 1986, p. 2-96.

Chapter 37

Fig. 37-1. Electronic Design, 3/75, p. 68.
Fig. 37-2. EXAR, Telecommunications Databook, 1986, p. 11-38.

Chapter 38

Fig. 38-1. Electronic Design, 8/73, p. 86.
Fig. 38-2. Electronic Design, 12/78, p. 98.
Fig. 38-3. Motorola, Thyristor Device Data, SEries A, 1985, p. 1-6-53.
Fig. 38-4. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-59.

Chapter 39

Fig. 39-1. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S5-143.

Chapter 40

Fig. 40-1. Electronic Design, 6/79, p. 122.
Fig. 40-2. Electronic Engineering, 9/84, p. 37.
Fig. 40-3. NASA Tech Briefs, 6/87, p. 26.
Fig. 40-4. Electronics Today International, 6/80, p. 68.
Fig. 40-5. Electronic Engineering, 9/87, p. 27.
Fig. 40-6. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 7-25.
Fig. 40-7. Radio-Electronics, 5/70, p. 33.
Fig. 40-8. Linear Technology Corp., Linear Databook, 1986, p. 5-78.
Fig. 40-9. Linear Technology Corp., Linear Databook, 1986, p. 8-40.
Fig. 40-10. Electronic Engineering, 2/79, p. 23.
Fig. 40-11. Electronic Engineering, 7/86, p. 30.
Fig. 40-12. Motorola, Application Note AN-294, p. 6.
Fig. 40-13. Motorola, Linear Integrated Circuits, p. 3-139.
Fig. 40-14. Hands-On Electronics, Winter 1985, p. 60.
Fig. 40-15. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 7-16.
Fig. 40-16. Signetics, Analog Data Manual, 1982, p. 3-39.
Fig. 40-17. General Electric/RCA, BiMOS Operational Amplifier Circuit Ideas, 1987, p. 10.
Fig. 40-18. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-20.
Fig. 40-19. National Semiconductor, Linear Brief 23.

Chapter 41

Fig. 41-1. Tab Books, Inc., 101 Sound, Light, and Power IC Projects.
Fig. 41-2. Courtesy, William Sheets.

Chapter 42

Fig. 42-1. Courtesy, William Sheets.
Fig. 42-2. Hands-On Electronics, Sep/Oct 1986, p. 85.
Fig. 42-3. Courtesy, William Sheets.

Chapter 43

Fig. 43-1. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 12-8.

Fig. 43-2. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 12-10.

Chapter 44

Fig. 44-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-7.

Chapter 45

Fig. 45-1. Linear Technology Corp., Linear Databook Supplement, 1988, p. S2-34.

Fig. 45-2. Hands-On Electronics, Jul/Aug 1986, p. 86.

Fig. 45-3. Signetics, Linear Data Manual Vol. 3: Video, p. 5-15.

Fig. 45-4. Tab Books, Inc. Build Your Own Laser, Phaser, Ion Ray Gun, 1983, p. 29.

Fig. 45-5. Hands-On Electronics, Jul/Aug 1986, p. 86.

Fig. 45-6. Electronic Engineering, 8/78, p. 24.

Chapter 46

Fig. 46-1. Hands-On Electronics, 12/86, p. 42.

Fig. 46-2. Linear Technology Corp., Linear Databook, 1986, p. 2-82.

Fig. 46-3. Electronic Engineering, 9/84, p. 33.

Fig. 46-4. Texas Instruments, Linear and Interface Circuits Applications Vol. 1, 1985, p. 3-18.

Fig. 46-5. Linear Technology Corp., Linear Databook, 1986, p. 2-83.

Chapter 47

Fig. 47-1. Electronic Engineering, 7/86, p. 30.

Fig. 47-2. Signetics, Analog Data Manual, 1982, p. 3-73.

Fig. 47-3.National Semiconductor Corp., Data Conversion/Acquisition Databook, 1980, p. 3-30.

Chapter 48

Fig. 48-1. Hands-On Electronics, 5/87, p. 95.

Fig. 48-2. Electronic Design 16, 8/76, p. 76.

Chapter 49

Fig. 49-1. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-50.

Fig. 49-2. Motorola, TMOS Power FET Design Ideas, 1985, p. 20.

Fig. 49-3. Motorola, TMOS Power FET Design Ideas, 1985, p. 21.

Fig. 49-4. R-E Experimenters Handbook, p. 156.

Fig. 49-5. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-48.

Fig. 49-6. Motorola, Thyristor Device Data, Series A, 1985, 1-6-55.

Fig. 49-7. Electronic Engineering, 9/84, p. 38.

Fig. 49-8. Tab Books, Inc., 101 Sound, Light, and Power IC Projects.

Fig. 49-9. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-60.

Fig. 49-10. General Electric, Application Note 200.35, p. 17.

Chapter 50

Fig. 50-1. Tab Books, Inc., Build Your Own Laser, Phasor, Ion Ray Gun, 1983, p. 104.

Fig. 50-2. Electric Engineering, 12/84, p. 34.

Chapter 51

Fig. 51-1. Hands-On Electronics, Sep/Oct 1986, p. 26.

Fig. 51-2. General Electric, Optoelectronics, Third Edition, p. 107.

Fig. 51-3. Courtesy, William Sheets.

Fig. 51-4. Courtesy, William Sheets.

Fig. 51-5. Electronic Engineering, 12/75, p. 15.

Fig. 51-6. Hands-On Electronics, 4/87, p. 94.

Fig. 51-7. Hands-On Electronics, 2/87, p. 87.

Fig. 51-8. Hands-On Electronics, 10/87, p. 92.

Fig. 51-9. Linear Technology Corp., Linear Databook, 1986, p. 2-83.

Fig. 51-10. Radio-Electronics, 11/86, p. 38.

Fig. 51-11. General Electric, Application Note 200.35, p. 15.

Fig. 51-12. Intersil, Component Data Catalog, 1987, p. 7-44.

Fig. 51-13. Radio Electronics, 3/86, p. 32.

Fig. 51-14. Electronic Design, 11/82, p. 172.

Fig. 51-15. Electronic Design, 6/76, p. 120.

Fig. 51-16. Linear Technology Corp., Linear Application Handbook, 1987, p. AN5-3.

Chapter 52

Fig. 52-1. Siliconix, Integrated Circuit Data Book, 3/85, p. 10-85.

Fig. 52-2. Siliconix, Integrated Circuit Data Book, 3/85, p. 10-79.
Fig. 52-3. Siliconix, Integrated Circuit Data Book, 3/85, p. 2-144.
Fig. 52-4. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 10/86, p. 4-261.
Fig. 52-5. Siliconix, Integrated Circuit Data Book, 3/85, p. 2-103.

Chapter 53

Fig. 53-1. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 2/87, p. 5-350.
Fig. 53-2. Linear Technology Corp., Linear Application Handbook, 1987, p. AN3-9.

Chapter 54

Fig. 54-1. Hands-On Electronics, May/Jun 1986, p. 63.
Fig. 54-2. Electronic Design, 10/73, p. 114.
Fig. 54-3. Electronic Engineering, 7/85, p. 44.
Fig. 54-4. Electronics Today International, 3/80, p. 25.
Fig. 54-5. Popular Electronics, 1/82, p. 76.
Fig. 54-6. Electronic Engineering, 6/87, p. 28.
Fig. 54-7. Popular Electronics, 8/69, p. 71.
Fig. 54-8. Electronics Today International, 1/76, p. 52.
Fig. 54-9. Electronic Engineering, 9/78, p. 20.

Chapter 55

Fig. 55-1. Electronic Engineering, 1/85, p. 39.
Fig. 55-2. Intersil, Component Data Catalog, 1987, p. 7-44.

Chapter 56

Fig. 56-1. Courtesy, William Sheets.
Fig. 56-2. Courtesy, William Sheets.

Chapter 57

Fig. 57-1. Popular Electronics, 6/73.
Fig. 57-2. Courtesy, William Sheets.
Fig. 57-3. Courtesy, William Sheets.
Fig. 57-4. Courtesy, William Sheets.

Chapter 58

Fig. 58-1. Unitrode Corp., 10/86, p. 332.
Fig. 58-2. Unitrode Corp., 10/86, p. 332.
Fig. 58-3. Radio Electronics, 8/82, p. 36.
Fig. 58-4. Hands-On Electronics, Winter 1985, p. 93.
Fig. 58-5. Hands-On Electronics, 9/87, p. 71.
Fig. 58-6. Electronic Engineering, 10/77, p. 23.
Fig. 58-7. Electronic Design, 11/8/69, p. 109.
Fig. 58-8. Radio-Electronics, 11/82, p. 79.
Fig. 58-9. National Semiconductor Corp., Transistor Databook, 1982, p. 11-34.
Fig. 58-10. National Semiconductor Corp., Data Conversion/Acquisition Databook, 1980, p. 2-5.
Fig. 58-11. Linear Technology Corp., Linear Databook, 1986, p. 8-42.
Fig. 58-12. Signetics, Linear Data Manual Vol. 3: Video, p. 11-120.
Fig. 58-13. General Electric, Application Note 90.16, p. 28.
Fig. 58-14. RCA, Digital Integrated Circuits Application Note ICAN-6346, p. 5.
Fig. 58-15. Linear Technology Corp., Linear Databook, 1986, p. 5-15.
Fig. 58-16. General Electric, SCR Manual, Sixth Edition, 1979, p. 204.

Chapter 59

Fig. 59-1. Motorola, TMOS Power FET Design Ideas, 1985, p. 45.
Fig. 59-2. Courtesy, William Sheets.
Fig. 59-3. Signetics, Linear Data Manual, Vol. 3: Video, p. 11-3.
Fig. 59-4. Radio-Electronics, 8/77, p. 33.
Fig. 59-5. Electronic Design, 3/77, p. 76.

Chapter 60

Fig. 60-1. Electronic Engineering, 5/84, p. 43.
Fig. 60-2. General Electric, Optoelectronics, Third Edition, p. 114.
Fig. 60-3. Motorola, TMOS Power FET Design Ideas, 1985, p. 32.
Fig. 60-4. Siliconix, MOSpower Applications Handbook, p. 6-186.
Fig. 60-5. Electronic Engineering, 7/86, p. 34.
Fig. 60-6. Electronic Engineering, 4/85, p. 47.
Fig. 60-7. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-8.
Fig. 60-8. Motorola, TMOS Power FET Design Ideas, 1985, p. 31.
Fig. 60-9. Sprague Electric Co., Integrated Circuits Databook WR504, p. 4-159.
Fig. 60-10. General Electric, Application Note 200.35, p. 18.
Fig. 60-11. Sprague Electric Co., Integrated Circuits Databook WR504, p. 4-160.

Fig. 60-12. Motorola, TMOS Power FET Design Ideas, 1985, p. 55.
Fig. 60-13. Motorola, TMOS Power FET Design Ideas, 1985, p. 54.
Fig. 60-14. Motorola, TMOS Power FET Design Ideas, 1985, p. 51.
Fig. 60-15. Electronic Engineering, 2/84, p. 23.
Fig. 60-16. National Semiconductor Corp., Linear Application Databook, p. 1066.
Fig. 60-17. General Electric, Optoelectronics, Third Edition, p. 113.

Chapter 61

Fig. 61-1. Fairchild Corp., Linear Databook, 1982, p. 4-72.
Fig. 61-2. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-14.

Chapter 62

Fig. 62-1. 73 Magazine, 12/76, p. 170.
Fig. 62-2. Electronics International Today, 1/76, p. 44.
Fig. 62-3. Signetics Analog Data Manual, 1983, p. 10-93.
Fig. 62-4. Electronics Today International, 9/75, p. 66.
Fig. 62-5. CQ, 5/76, p. 26.

Chapter 63

Fig. 63-1. National Semiconductor Corp., Linear Applications Databook, p. 1096.
Fig. 63-2. EXAR, Telecommunications Databook, 1986, p. 7-24.
Fig. 63-3. EXAR, Telecommunications Databook, 1986, p. 7-24.
Fig. 63-4. Electronic Engineering, 12/84, p. 33.
Fig. 63-5. Electronic Engineering, 11/85, p. 31.
Fig. 63-6. Courtesy, William Sheets.
Fig. 63-7. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 2-11.
Fig. 63-8. Courtesy, William Sheets.
Fig. 63-9. Courtesy, William Sheets.
Fig. 63-10. Courtesy, William Sheets.

Chapter 64

Fig. 64-1. Electronic Engineering, 6/83, p. 31.
Fig. 64-2. Electronic Design 15, 7/75, p. 68.

Chapter 65

Fig. 65-1. General Electric, Optoelectronics, Third Edition, p. 135.
Fig. 65-2. NASA, Tech Briefs, Summer 1984, p. 446.
Fig. 65-3. General Electric, Optoelectronics, Third Edition, p. 140.
Fig. 65-4. General Electric, Optoelectronics, Third Edition, p. 121.
Fig. 65-5. General Electric, Optoelectronics, Third Edition, p. 120.
Fig. 65-6. General Electric, Optoelectronics, Third Edition, p. 139.
Fig. 65-7. General Electric, Optoelectronics, Third Edition, p. 120.
Fig. 65-8. General Electric, Optoelectronics, Third Edition, p. 134.
Fig. 65-9. General Electric, Optoelectronics, Third Edition, p. 112.
Fig. 65-10. National Semiconductor Corp., Data Conversion/Acquisition Databook, 1980, p. 13-46.
Fig. 65-11. General Electric, Optoelectronics, Third Edition, p. 133.
Fig. 65-12. Courtesy, William Sheets.
Fig. 65-13. General Electric, Optoelectronics, Third Edition, p. 117.
Fig. 65-14. Electronic Engineering, 8/86, p. 36.
Fig. 65-15. General Electric, Optoelectronics, Third Edition, p. 118.

Chapter 66

Fig. 66-1. 73 For Radio Amateurs, 11/85, p. 32.
Fig. 66-2. Radio-Electronics, 5/70, p. 35.
Fig. 66-3. Electronics Today International, 7/78, p. 16.
Fig. 66-4. Electronics Today International, 12/78, p. 15.
Fig. 66-5. General Electric, Semiconductor Data Handbook, Third Edition, p. 513.
Fig. 66-6. Electronic Design, 11/29/84, p. 281.
Fig. 66-7. Unitrode Corp., Databook 1986, p. 51.
Fig. 66-8. Electronic Engineering, 5/77, p. 27.
Fig. 66-9. National Semiconductor Corp., Transistor Databook, 1982, p. 7-19.
Fig. 66-10. Courtesy, William Sheets.
Fig. 66-11. Electronic Design, 10/65.

Fig. 66-12. Hands-On Electronics, Summer 1984, p. 43.
Fig. 66-13. Signetics, Analog Data Manual, 1982, p. 8-10.

Chapter 67

Fig. 67-1. Electronic Design, 5/79, p. 102.
Fig. 67-2. Radio-Electronics, 7/70, p. 36.
Fig. 67-3. Courtesy, William Sheets.

Chapter 68

Fig. 68-1. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-18.
Fig. 68-2. Popular Electronics, 3/79, p. 78.

Chapter 69

Fig. 69-1. Electronic Engineering, 2/86, p. 38.
Fig. 69-2. Courtesy, William Sheets.
Fig. 69-3. Electronic Engineering, 4/77, p. 13.
Fig. 69-4. Electronic Engineering, 7/85, p. 34.
Fig. 69-5. Electronic Design, 3/77, p. 106.
Fig. 69-6. Electric Engineering, 1/87, p. 25.

Chapter 70

Fig. 70-1. Hands-On Electronics, 10/87, p. 96.
Fig. 70-2. Hands-On Electronics, Spring 1985, p. 82.
Fig. 70-3. General Electric Project G4, p. 131.
Fig. 70-4. Radio Electronics, 12/84, p. 77.
Fig. 70-5. Electronics Today International, 6/75, p. 42.
Fig. 70-6. Electronics Today International, 9/82, p. 42.

Chapter 71

Fig. 71-1. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/86, p. 7-251.
Fig. 71-2. National Semiconductor Corp., Linear Applications Databook, p. 1065.
Fig. 71-3. Siliconix, MOSpower Applications Handbook, p. 6-101.
Fig. 71-4. Hands-On Electronics, 5/87, p. 96.
Fig. 71-5. Hands-On Electronics, Spring 1985, p. 36.
Fig. 71-6. Hands-On Electronics, Summer 1984, p. 74.
Fig. 71-7. Radio-Electronics, 3/86, p. 59.
Fig. 71-8. National Semiconductor Corp., Audio/Radio Handbook, 1980, p. 4-20.
Fig. 71-9. National Semiconductor Corp., Linear Databook, 1982, p. 3-187.
Fig. 71-10. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 11/86, p. 4-135.

Chapter 72

Fig. 72-1. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, , p. 6-35.
Fig. 72-2. Motorola, TMOS Power FET Design Ideas, 1985, p. 43.
Fig. 72-3. Electronic Engineering, 12/84, p. 41.
Fig. 72-4. NASA, Tech Briefs, 9/87, p. 21.
Fig. 72-5. Motorola, TMOS Power FET Design Ideas, 1985, p. 37.
Fig. 72-6. Siliconix, MOSpower Applications Handbook, p. 6-51.
Fig. 72-7. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 24.
Fig. 72-8. Courtesy, William Sheets.
Fig. 72-9. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 24.
Fig. 72-10. Radio-Electronics, 6/86, p. 52.
Fig. 72-11. Siliconix, MOSpower Applications Handbook, p. 6-177.
Fig. 72-12. Siliconix, MOSpower Applications Handbook, p. 6-59.
Fig. 72-13. Linear Technology Corp., Linear Databook, 1986, p. 3-23.
Fig. 72-14. Electronic Engineering, 10/84, p. 38.
Fig. 72-15. NASA Tech Briefs, Summer 1985, p. 32.
Fig. 72-16. Electronic Engineering, 1/87, p. 22.
Fig. 72-17. Motorola, TMOS Power FET Design Ideas, 1985, p. 42.
Fig. 72-18. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-55.
Fig. 72-19. 73 Magazine, 12/70, p. 170.
Fig. 72-20. Signetics, Analog Data Manual, 1983, p. 12-27.
Fig. 72-21. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 2/87, p. 8-223.
Fig. 72-22. Electronic Engineering, 1/85, p. 45.
Fig. 72-23. Motorola, Linear Integrated Circuits, p. 3-138.

Fig. 72-24. Electronic Design, 11/29/84, p. 282.

Fig. 72-25. Electronics Today International, 1/70, p. 45.

Fig. 72-26. Linear Technology, 1986 Linear Databook, p. 3-22.

Chapter 73

Fig. 73-1. Electronic Engineering, 10/76, p.17.

Fig. 73-2. Electronic Engineering, 7/77, p. 26.

Fig. 73-3. Popular Electronics, 5/74, p. 24.

Fig. 73-4. Courtesy, William Sheets.

Chapter 74

Fig. 74-1. NASA, Tech Briefs, Winter 1985, p. 52.

Fig. 74-2. Courtesy, William Sheets.

Fig. 74-3. Electronic Engineering, 3/86, p. 34.

Fig. 74-4. Courtesy, William Sheets.

Fig. 74-5. Courtesy, William Sheets.

Chapter 75

Fig. 75-1. Electronic Design 25, 1275, p. 90.

Fig. 75-2. Courtesy, William Sheets.

Fig. 75-3. Electronics Today International, 9/75, p. 66.

Fig. 75-4. Electronic Engineering, 1/85, p. 41.

Fig. 75-5. Hands-On Electronics, Fall 1984, p. 66.

Fig. 75-6. Radio-Electronics, 3/77, p. 76.

Chapter 76

Fig. 76-1. Hands-On Electronics, 11/86, p. 92.

Fig. 76-2. Popular Electronics, 11/77, p. 62.

Chapter 77

Fig. 77-1. Electronic Engineering, 9/86, p. 38.

Fig. 77-2. RCA, Design Guide for Fire Detection Systems, Publication 2M1189, p. 27.

Fig. 77-3. Electronic Engineering, 9/86, p. 34.

Chapter 78

Fig. 78-1. Electronic Engineering, 5/76, p. 17.

Fig. 78-2. Electronic Design, 4/74, p. 114.

Fig. 78-3. Electronic Engineering, 10/86, p. 41.

Chapter 79

Fig. 79-1. Electronic Engineering, 12/75, p. 15.

Fig. 79-2. Radio-Electronics, Experimenters Handbook, p. 122.

Chapter 80

Fig. 80-1. Signetics, 1987 Linear Data Manual, Vol. 1: Communications, 2/87, p. 4-310.

Fig. 80-2.Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-52.

Chapter 81

Fig. 81-1.Signetics, 1987 Linear Data Manual, Vol. 1: Communications, 11/86, p. 7-14.

Fig. 81-2. Hands-On Electronics, 3/87, p. 28.

Fig. 81-3. Hands-On Electronics, 12/86, p. 22.

Chapter 82

Fig. 82-1. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 11.

Fig. 82-2. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 11/86, p. 4-135.

Chapter 83

Fig. 83-1. Electronic Design, 9/69, p. 106.

Fig. 83-2. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-61.

Fig. 83-3. QST, 7/87, p. 32.

Chapter 84

Fig. 84-1. Electronic Engineering, 4/86, p. 34.

Fig. 84-2. Electronic Engineering, 7/85, p. 44.

Fig. 84-3. Electronic Engineering, 11/86, p. 34.

Fig. 84-4. NASA, Tech Briefs, 1/88, p. 18.

Chapter 85

Fig. 85-1. Motorola, RF Data Manual, 1986, p. 6-141.

Fig. 85-2. Motorola, RF Data Manual, 1986, p. 6-240.

Fig. 85-3. QST, 7/87, p. 31.

Fig. 85-4. QST, 5-86, p. 23.

Fig. 85-5. Motorola, RF Data Manual, 1986, p. 6-181.

Fig. 85-6. Ham Radio, 7/86, p. 50.

Fig. 85-7. Radio Electronics, 3/87, p. 42.

Fig. 85-8. NASA, Tech Briefs, Spring 1984, p. 322.

Fig. 85-9. Motorola, RF Data Manual, 1986, p. 6-232.

Chapter 86

Fig. 86-1. QST, 12/85, p. 39.

Chapter 87

Fig. 87-1. General Electric/RCA, BiMOS Operational Amplifiers circuit Ideas, 1987, p. 14.
Fig. 87-2. Intersil, Component Data Catalog, 1987, p. 7-5.
Fig. 87-3. Linear Technology Corp., Linear Databook, 1986, p. 2-113.
Fig. 87-4. Electronics Today International, 3/78, p. 51.
Fig. 87-5. National Semiconductor Corp., Hybrid Products Databook, 1982, p. 17-149.
Fig. 87-6. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-58.

Chapter 88

Fig. 88-1. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 8.
Fig. 88-2. Linear Technology Corp., Linear Databook, 1986, p. 2-113.
Fig. 88-3. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-15, 3-16.
Fig. 88-4. QST, 4/87, p. 48.
Fig. 88-5. Electronic Engineering, 5/85, p. 38.
Fig. 88-6. Electronic Design, 6/81, p. 250.
Fig. 88-7. Ham Radio, 1/87, p. 97.
Fig. 88-8. Electronic Engineering, 2/76, p. 17.
Fig. 88-9. Electronic Design, 2/73, p. 82.
Fig. 88-10. Radio-Electronics, 2/71, p. 37.
Fig. 88-11. Ham Radio, 6/82, p. 33.

Chapter 89

Fig. 89-1. Popular Electronics, 12/74, p. 68.
Fig. 89-2. Electronics Today International, 1/77, p. 49.
Fig. 89-3. Electronics Today International, 1/77, p. 49.
Fig. 89-4. Popular Electronics, 8/74, p. 98.
Fig. 89-5. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 28.
Fig. 89-6. Electronics Today International, 11/76, p. 45.
Fig. 89-7. Electronics Today International, 2/75, p. 66.
Fig. 89-8. Electronics Today International, 1/77, p. 85.
Fig. 89-9. Electronics Today International, 6/75, p. 63.
Fig. 89-10. Electronics Today International, 1/77, p. 49.
Fig. 89-11. Courtesy, William Sheets.
Fig. 89-12. Electronics Today International, 11/80, p. 43.
Fig. 89-13. Radio-Electronics, 2/75, p. 42.
Fig. 89-14. Courtesy, William Sheets.

Chapter 90

Fig. 90-1. Courtesy, William Sheets.
Fig. 90-2. Radio-Electronics, 12/83, p. 38.
Fig. 90-3. Hands-On Electronics, 8/87, p. 77.
Fig. 90-4. Electronics today International, 6/79, p. 27.
Fig. 90-5. Courtesy, William Sheets.

Chapter 91

Fig. 91-1. Hands-On Electronics, 6/87, p. 40.
Fig. 91-2. Electronics Today International, 8/77, p. 25.
Fig. 91-3. Electronics Today International, 11/80.
Fig. 91-4. Texas Instruments, Complex Sound Generator, Bulletin No. DL-12612, p. 13.
Fig. 91-5. Electronics Today International, 4/82, p. 34.
Fig. 91-6. Electronics Today International, 2/75, p. 66.
Fig. 91-7. Hands-On Electronics, 12/86, p. 42.
Fig. 91-8. Courtesy, William Sheets.
Fig. 91-9. Courtesy, William Sheets.
Fig. 91-10. Texas Instruments, complex Sound Generator, Bulletin No. DL-S 12612, p. 11.
Fig. 91-11. Courtesy, William Sheets.

Chapter 92

Fig. 92-1. Electronic Engineering, 3/82, p. 29.
Fig. 92-2. Electronic Engineering, 1078, p. 17.
Fig. 92-3. EXAR, Telecommunications Databook, 1986, p. 9-24.
Fig. 92-4. Electronic Design, 5/79, p. 100.
Fig. 92-5. Siliconix, Integrated Circuits Data Book, 3/85, p. 5-17.
Fig. 92-6. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 7.

Fig. 92-7. Electronic Design, 6/69, p. 126.
Fig. 92-8. Electronic Engineering, 8/84, p. 27.
Fig. 92-9. Electronic Engineering, 12/85, p. 35.
Fig. 92-10. Electronic Engineering, 8/84, p. 29.

Chapter 93

Fig. 93-1. General Electric/RCA, Operational Amplifiers Circuit Ideas, 1987, p. 10.
Fig. 93-2. Linear Technology Corp., Linear Databook, 1986, p. 8-42.

Chapter 94

Fig. 94-1. Tab Books, Inc. 303 Dynamic Electronic Circuits, p. 169.
Fig. 94-2. Electronics Today International, 12/77, p. 86.

Chapter 95

Fig. 95-1. Electronics Today International, 10/78, p. 46.
Fig. 95-2. Hands-On Electronics, Fall 1984, p. 65.
Fig. 95-3. Courtesy, William Sheets.

Chapter 96

Fig. 96-1. Electronic Engineering, 6/86, p. 35.
Fig. 96-2. General Electric, SCR Manual, Sixth Edition, 1979, p. 200.

Chapter 97

Fig. 97-1. Electronic Engineering, 9/87, p. 32.
Fig. 97-2. Intersil, Component Data Catalog, 1987, p. 14-67.

Chapter 98

Fig. 98-1. Electronic Engineering, 2/87, p. 40.
Fig. 98-2. Tab Books, Inc. The Giant Book of Easy-To-Build Electronic Projects, 1982, p. 1.
Fig. 98-3. Radio-Electronics, 2/85, p. 90.
Fig. 98-4. EXAR, Telecommunications Databook, 1986, p. 11-38.
Fig. 98-5. Tab Books, Inc. build Your Own Laser, Phaser, Ion gun, 1983, p. 305.
Fig. 98-6. Electronic Design, 12/78, p. 95.
Fig. 98-7. Radio-Electronics, 11/79, p. 53.
Fig. 98-8. Electronic Design, 10/76, p. 194.
Fig. 98-9. Hands-On Electronics, 12/86, p. 22.
Fig. 98-10. Electronics Engineering, 1/79, p. 17.
Fig. 98-11. Radio-Electronics, 12/78, p. 67.
Fig. 98-12. Radio-Electronics, 11/77, p. 45.
Fig. 98-13. Hands-On Electronics, Summer 1985, p. 74.
Fig. 98-14. Hands-On Electronics, Sep/Oct 1986, p. 88.
Fig. 98-15. Hands-On Electronics, Sep/Oct 1986, p. 105.
Fig. 98-16. Hands-On Electronics, Summer 1984, p. 39.
Fig. 98-17. EXAR, Telecommunications Databook, 1986, p. 4-19.
Fig. 98-18. EXAR, Telecommunications Databook, 1986, p. 5-14.
Fig. 98-19. EXAR, Telecommunications Databook, 1986, p. 4-15.
Fig. 98-20. Signetics, 1987 Linear Data Manual, Vol. 1: Communications, 12/2/86, p. 6-50.

Chapter 99

Fig. 99-1. NASA, Tech Briefs, 12/87, p. 28.
Fig. 99-2. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 2/87, p. 7-67.
Fig. 99-3. Electronic Design, 8/75, p. 82.
Fig. 99-4. Linear Technology Corp., Linear Application Handbook, 1987, p. AN3-6.
Fig. 99-5. General Electric, Optoelectronics, Third Edition, p. 153.
Fig. 99-6. General Electric, Application Note 200.85, p. 18.
Fig. 99-7. Electronic Design, 8/82, p. 217.
Fig. 99-8. Electronic Design, 8/83, p. 230.
Fig. 99-9. Courtesy, William Sheets.

Chapter 100

Fig. 100-1. General Electric, SCR Manual, Sixth Edition, 1979, p. 222.
Fig. 100-2. Electronic Engineering, 9/85, p. 30.
Fig. 100-3. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S1-41.
Fig. 100-4. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S1-42.
Fig. 100-5. Linear Technology Corp., Linear Databook, 1986, p. 2-101.
Fig. 100-6. Courtesy, William Sheets.

Chapter 101

Fig. 101-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN7-2.
Fig. 101-2. Electric Engineering, 7/84, p. 31.

Chapter 102

Fig. 102-1. Hands-On Electronics, 11/86, p. 93.
Fig. 102-2. Hands-On Electronics, 9/87, p. 32.

Chapter 103

Fig. 103-1. Teledyne Semiconductor, Data & Design Manual, 1981, p. 7-17.
Fig. 103-2. Linear Technology Corp., Application Note 9, p. 18.
Fig. 103-3. Radio-Electronics, 9/82, p. 42.
Fig. 103-4. Intersil, Component Data Catalog, 1987, p. 6-8.
Fig. 103-5. Intersil, Component Data Catalog, 1987, p. 6-11.
Fig. 103-6. Intersil, Component Data Catalog, 1987, p. 6-8.

Chapter 104

Fig. 104-1. NASA, Tech Briefs, Spring 1985, p. 40.
Fig. 104-2. NASA, Tech Briefs, Fall/Winter 1981, p. 319.
Fig. 104-3. Popular Electronics, 12/82, p. 82.

Chapter 105

Fig. 105-1. General Electric, Semiconductor Data Handbook, Third Edition, p. 513.
Fig. 105-2. Motorola, Application Note AN294.
Fig. 105-3. Electronic Design, 4/77, p. 120.
Fig. 105-4. Motorola, Application Note AN294.

Chapter 106

Fig. 106-1. NASA, Tech Briefs, Sep/Oct 1986, p. 36.
Fig. 106-2. Electronic Engineering, 9/77, p. 37.
Fig. 106-3. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-51.
Fig. 106-4. Motorola, Thyristor Device Data, Series A, 1985, p. 1-6-54.
Fig. 106-5. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 7-23.
Fig. 106-6. Intersil, Databook, 1987, p. 7-102.
Fig. 106-7. Intersil, Component Data Catalog, 1987, p. 14-67.
Fig. 106-8. General Electric, Application Note 90.16, p. 30.

Chapter 107

Fig. 107-1. Electronics Today International, 11/80.
Fig. 107-2. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 2/87, p. 4-107.
Fig. 107-3. Electronics Today International, 9/77, p. 55.
Fig. 107-4. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 21.
Fig. 107-5. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 3-11.
Fig. 107-6. National Semiconductor Corp., Transistor Databook, 1982, p. 11-35.
Fig. 107-7. Electronics Today International, 6/82, p. 61.

Chapter 108

Fig. 108-1. Hands-On Electronics, 9/87, p. 88.
Fig. 108-2. Texas Instruments, Linear and Interface Circuits Applications, Vol. 1, 1985, p. 7-15.
Fig. 108-3. Tab Books, Inc., The Giant book of Easy-To-Build Electronics Projects, 1982, p. 31.

Index

Numbers preceded by a "I," "II," "III," or "IV" are from *Encyclopedia of Electronic Circuits* Vol. I, II, III, or IV, respectively.

B

C

D

E

F

G

H

I

J

K

L

M

N

O

P

Q

R

S

T

U

V

W

X

Y

Z

Other Bestsellers of Related Interest

ENCYCLOPEDIA OF ELECTRONIC CIRCUITS Vol. 1—Rudolf F. Graf

". . . schematics that encompass virtually the entire spectrum of electronics technology . . . This is a well worthwhile book to have handy." —***Modern Electronics***

Discover hundreds of the most versatile electronic and integrated circuit designs, all available at at the turn of a page. You'll find circuit diagrams and schematics for a wide variety of practical applications. Many entries also include clear, concise explanations of the circuit configurations and functions. 768 pages, 1,762 illustrations. Book No. 1938, $32.95 paperback, $60.00 hardcover

THE ILLUSTRATED DICTIONARY OF ELECTRONICS—5th Edition
—Rufus P. Turner and Stan Gibilisco

This completely revised and updated edition defines more than 27,000 practical electronics terms, acronyms, and abbreviations. Find up-to-date information on basic electronics, computers, mathematics, electricity, communications, and state-of-the-art applications—all discussed in a nontechnical style. The author also includes 360 new definitions and 125 illustrations and diagrams. 736 pages, 650 illustrations. Book No. 3345, $26.95 paperback

ELECTRONIC CONVERSIONS: Symbols and Formulas—2nd Edition
—Rufus P. Turner and Stan Gibilisco

This revised and updated edition supplies all the formulas, symbols, tables, and conversion factors commonly used in electronics. Exceptionally easy to use, the material is organized by subject matter. Its format is ideal and you can save time by directly accessing specific information. Topics cover only the most-needed facts about the most often used conversions, symbols, formulas, and tables. 280 pages, 94 illustrations. Book No. 2865, $14.95 paperback only

ELECTRONIC DATABOOK—4th Edition
—Rudolf F. Graf

If it's electronic, it's here—current, detailed, and comprehensive! Use this book to broaden your electronics information base. Revised and expanded to include all up-to-date information, this fourth edition makes any electronic job easier and less time-consuming. You'll find information that will aid in the design of local area networks, computer interfacing structure, and more! 528 pages, 131 illustrations. Book No. 2958, $24.95 paperback only

BUILD YOUR OWN TEST EQUIPMENT
—Homer L. Davidson

Build more than 30 common electronic testing devices, ranging from simple continuity and polarity testers to signal injectors and power supplies. Also learn how test instruments work, how they are used, and how to save money. Each project includes a complete parts list with exact part numbers. 300 pages, 324 illustrations. Book No. 3475, $17.95 paperback, $27.95 hardcover

ELECTRONIC SIGNALS AND SYSTEMS: Television, Stereo, Satellite TV, and Automotive—Stan Prentiss

Study signal analysis as it applies to the operation and signal-generating capabilities of today's most advanced electronic devices with this handbook. It explains the composition and use of a wide variety of test instruments, transmission media, satellite systems, stereo broadcast and reception facilities, antennas, television equipment, and even automotive electrical systems. You'll find coverage of C- and Ku-band video, satellite master TV systems, high-definition television, C-QUAM® AM stereo transmission and reception, and more. 328 pages, 186 illustrations. Book No. 3557, $19.95 paperback, $29.95 hardcover

Prices Subject to Change Without Notice.

Look for These and Other TAB Books at Your Local Bookstore

To Order Call Toll Free 1-800-822-8158

(in PA, AK, and Canada call 717-794-2191)

or write to TAB Books, Blue Ridge Summit, PA 17294-0840.

Title	Product No.	Quantity	Price

☐ Check or money order made payable to TAB Books

Charge my ☐ VISA ☐ MasterCard ☐ American Express

Acct. No. ______________ Exp. ________

Signature: ______________

Name: ______________

Address: ______________

City: ______________

State: ________ Zip: ________

Subtotal $ ________

Postage and Handling ($3.00 in U.S., $5.00 outside U.S.) $ ________

Add applicable state and local sales tax $ ________

TOTAL $ ________

TAB Books catalog free with purchase; otherwise send $1.00 in check or money order and receive $1.00 credit on your next purchase.

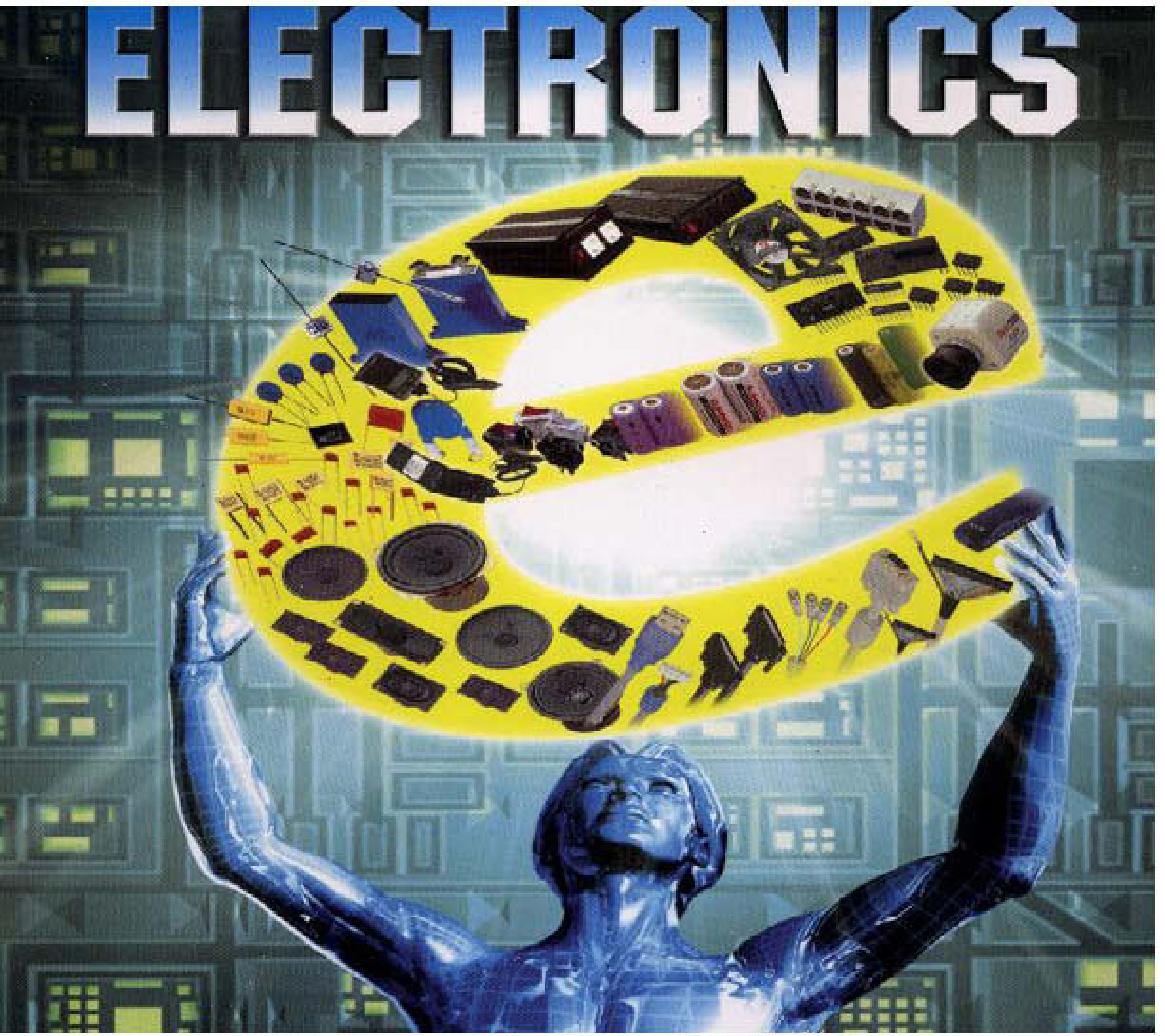
Orders outside U.S. must pay with international money in U.S. dollars

TAB Guarantee: If for any reason you are not satisfied with the book(s) you order, simply return it (them) within 15 days and receive a full refund. **BC**

Rudolf F. Graf

Encyclopedia of ELECTRONIC CIRCUITS

Volume 3


ELECTRONICS

Contents

To Sheryl Melissa,
a budding scholar
From Popsi

Preface

Volume III of *The Encyclopedia of Electronic Circuits* adds about 1,000 new circuits to the ready-to-use files that were established by the publication of volumes I and II of this set of circuits encyclopedias.

These three volumes now offer an invaluable storehouse of about 3,000 carefully arranged and categorized, easy-to-access circuits. Volume IV is scheduled for publication in 1992.

Once again it gives me great pleasure to extend my gratitude to William Sheets for his comments and contributions, and to Mrs. Stella Dillon for her virtuoso performance on the word processor.

1

Active Antennas

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

ACTIVE ANTENNA

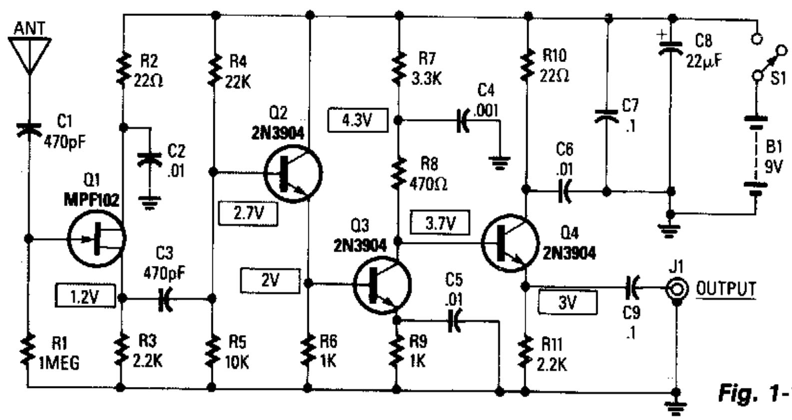


Fig. 1-1

Antennas that are much shorter than 1/4 wavelength present a very small and highly relative impedance that is dependent on the received frequency. It is difficult to match impedances over a decade of frequency coverage. Instead, input stage Q1 is an FET source-follower. A high-impedance input successfully bridges antenna characteristics at any frequency.

Transistor Q2 is used as an emitter-follower to provide a high-impedance load for Q1, but more importantly, it provides a low-drive impedance for common-emitter amplifier Q3, which provides all of the amplifier's voltage gain. Transistor Q4 transforms Q3's moderate output impedance into low impedance, thereby providing sufficient drive for a receiver's 50-Ω, antenna-input impedance.

ACTIVE ANTENNA

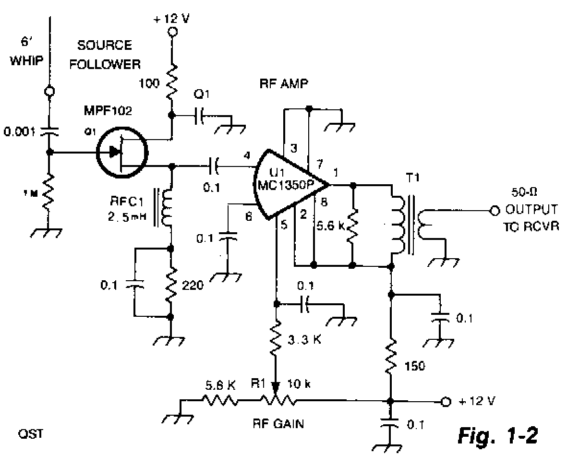


Fig. 1-2

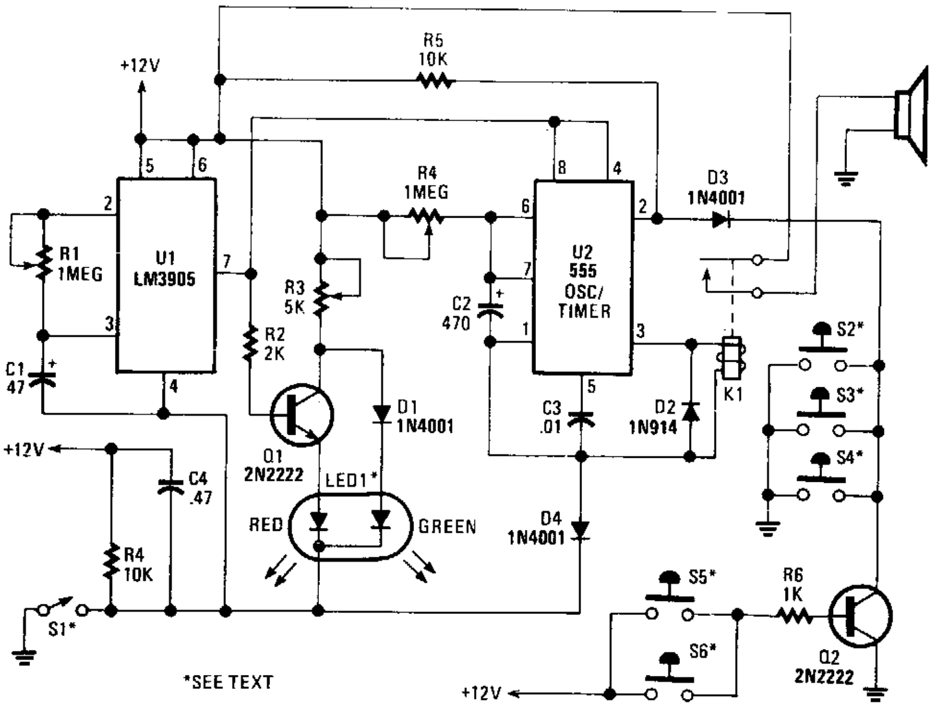
All capacitors in this circuit are disc ceramic. Fixed-value resistors are 1/4- or 1/2-W carbon. R1 controls the gain of U1. RFC1 is a miniature 2.5-mH rf choke. T1 has 30 primary turns of #28 enamel wire on an Amidon FT50 – 43 ferrite toroid core, and the secondary has four turns of #28 wire.

2

Alarm and Security Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

AUTO ALARM



POPULAR ELECTRONICS

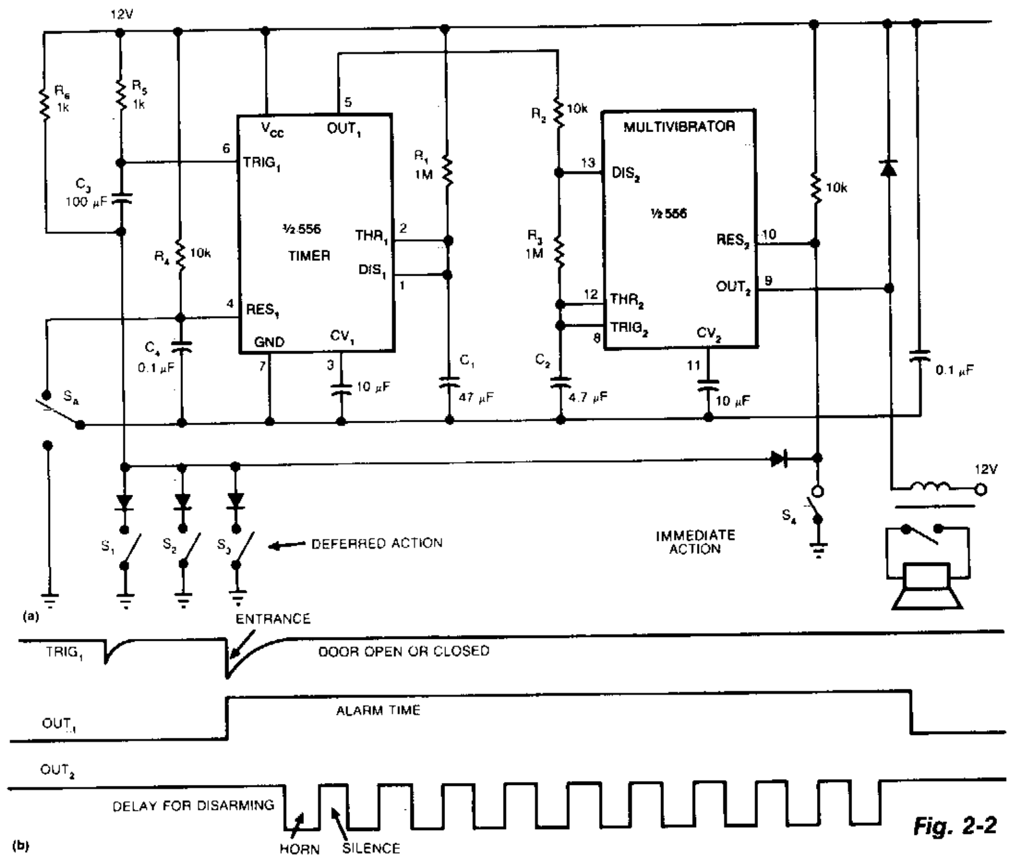
Fig. 2-1

In operation, the alarm circuit allows a 0 – 47 second time delay, as determined by the R1/C1 combination, after the switch is armed to allow the vehicle's motion sensor to settle down. This allows you time to get a bag of groceries out of the trunk and not have the hassle of juggling the groceries and the key switch at once.

During the time delay, half of LED1, which is actually a single, bi-colored, three-legged common cathode device, lights green. At the same time, pins 8 and 4 of U2 (a 555 oscillator/timer) are held low by U1 (a 3905 oscillator/timer), causing the alarm to remain silent. Once the delay is over, LED1 turns red, indicating that the circuit is armed.

At that point, a ground at pin 2 of U2 forces pin 3 of U2 high, closing the contacts of K1 and sounding the siren for a time duration determined by R4 and C2. Once the time has elapsed, pin 3 is pulled low, K1 opens, and the circuit is again ready to go. The circuit can be manually reset by the simple expedient of opening and closing the key switch. Potentiometer R3 controls the LED's illumination intensity. Diode D1 ensures that the green segment of LED1 is fully extinguished when Q1 is turned on – which turns the LED to red. Resistors R4 and R5 must be connected to the +V bus, not to pin 7 of U1, otherwise U2 will mysteriously trigger itself each time the initial delay ends.

ONE-CHIP BURGLAR ALARM

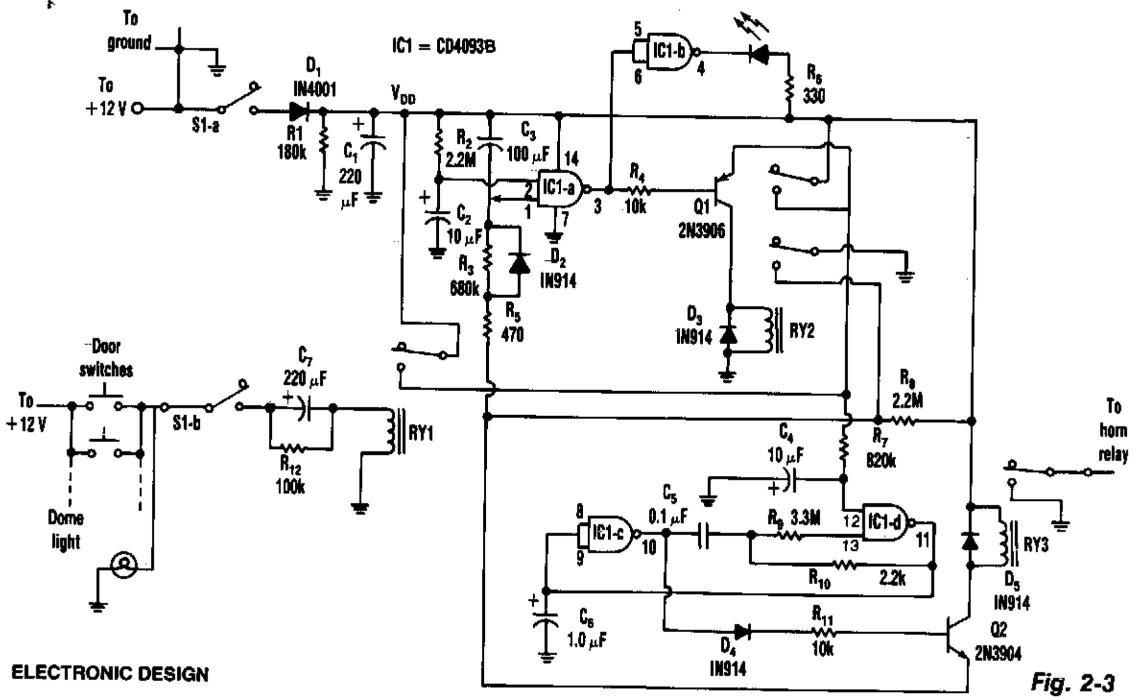


EDN

Fig. 2-2

The single-chip, burglar-alarm circuit shown uses a dual 556 timer, draws 10 mA of standby current, and generates a pulsing alarm signal that conserves battery energy. Once activated, the alarm will remain on, independent of the subsequent state of any of the sensors. The sensors support both deferred and immediate-action modes. You can attach this circuit to your car's internal lighting circuitry using a single wire and a relay. To arm this circuit, you open your car door and close switch SA. The switch discharges capacitor C4 and holds the timer (one half of the 556 IC) in a reset state to prevent false triggering while you're arming the circuit. When you close your car door, the circuit enters a standby mode. If the door is then reopened, the sensors apply a negative-going pulse to trigger 1. Output 1 then increases, enabling the alarm for 1.1R1C1 seconds. Output 1's high state triggers the multivibrator, the other half of the 556, which begins to cycle after a delay equal to 1.1 (R2 + R3) C2 seconds. As long as the timer's output stays high, the multivibrator will continue to cycle, turning the horn off and on at 3.3-second intervals. During the interval between time that the timer's output increases and the time that the multivibrator's output decreases, you can disarm this circuit using switch SA. To prevent false triggering caused by switch contacts, at S1, S2, and S3, that may bounce when closing the door, make the R6C3 time constant as large as possible. In addition, capacitors C1 and C2 should be tantalum types and should exhibit leakage of less than 1 μA at room temperature.

SEMICONDUCTOR FAIL-SAFE ALARM



ELECTRONIC DESIGN

Fig. 2-3

False alarms produced by semiconductor failure are impossible with this burglar-alarm circuit equipped with relays. What's more, the circuit is virtually immune to false triggering. With a standby current of less than 0.1 mA, the circuit offers all the features an alarm needs: entry and exit delays, a timed alarm period, and automatic reset after an intrusion.

One CMOS CD4093B quad NAND gate, IC1, supplies both logic and analog timing functions with the aid of Schmitt-trigger switching action. Relays make the circuit fail-safe in the alarm-active mode, even when the semiconductors fail. The relays are 12-V, with coil resistances of 250 Ω or more.

Closing switch S1 initiates circuit operation. Capacitor C2 begins charging through resistor R2 and arming indicator LED1 lights. When pin 2 of IC1a reaches its switching point, its output decreases, extinguishing LED1 and indicating that the exit delay has ended. That output also drives the base of Q1 low, so that if the emitter circuit completes to the V_{DD} line, Q1 conducts. The circuit is now armed, and current drain drops to less than 0.1 mA.

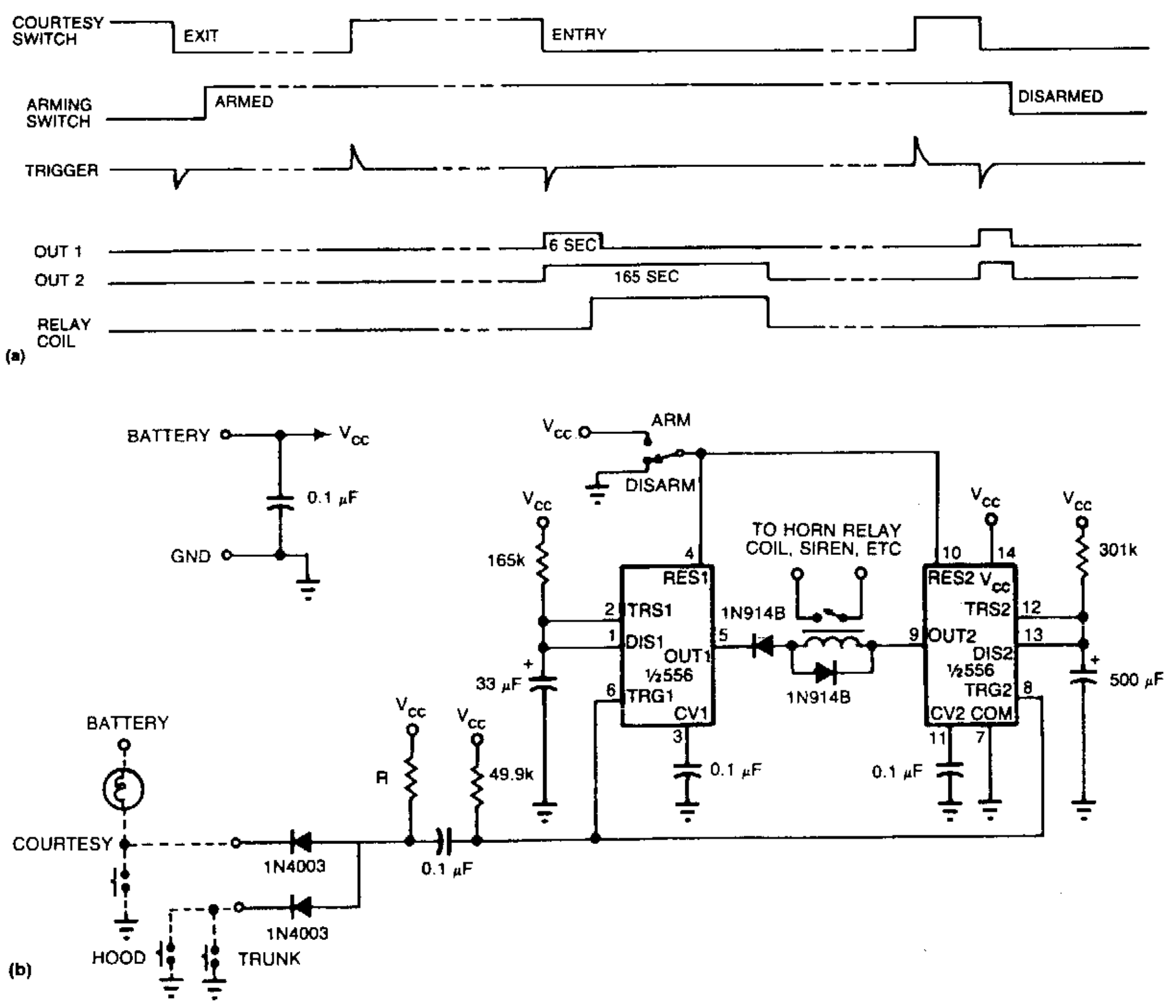
When the vehicle is entered, relay RY1 contacts close momentarily, completing the emitter circuit of Q1 and causing the RY2 contacts to close. Charging C4 through R7 determines the entry-delay period. If the system isn't turned off by opening S1 during this period, the oscillator circuit of IC1c and IC1d activates, and a rapid on/off horn-honking cycle kicks on with the aid of Q2 and RY3.

The alarm cycle ends after about a minute, when C2 charges through R3 to the threshold voltage of IC1a at pin 1. This voltage resets the timing circuit, readying it for another entry/alarm cycle.

RY1 is connected for vehicles that use door switches connected to +12 V. For vehicles that use grounding door switches, the bottom of the RY1 coil should connect to +12 V instead of ground. In the latter case, the polarity of C7 should be reversed.

For home use, the R3C3 time constant should be increased to give a longer alarm.

SINGLE-IC AUTO ALARM

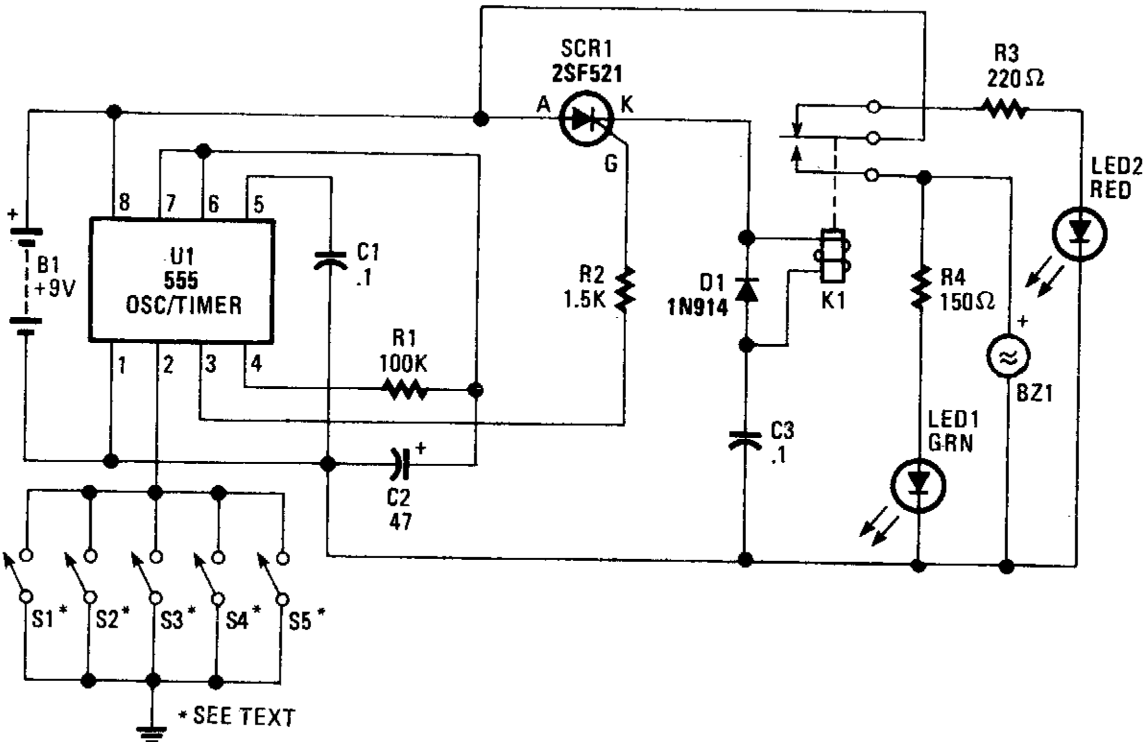


EDN

Fig. 2-4

See (a) for the timing information for the alarm circuit in (b). When leaving your vehicle, flip the arming switch and close the door to arm the device. Subsequent opening of an entrance triggers both timers. After the expiration of the entry delay timer, the alarm sounds for a time determined by the second timer. The value of *R* should be less than 1 KΩ. If you use an incandescent lamp instead of a resistor, you get an extra function—an open-entrance indicator. By keeping the resistance low, you avoid false tripping should water collect under the hood. If your door switch connects the courtesy light to 12 V rather than ground, use a single transistor as an inverter at the input.

BURGLAR ALARM



POPULAR ELECTRONICS

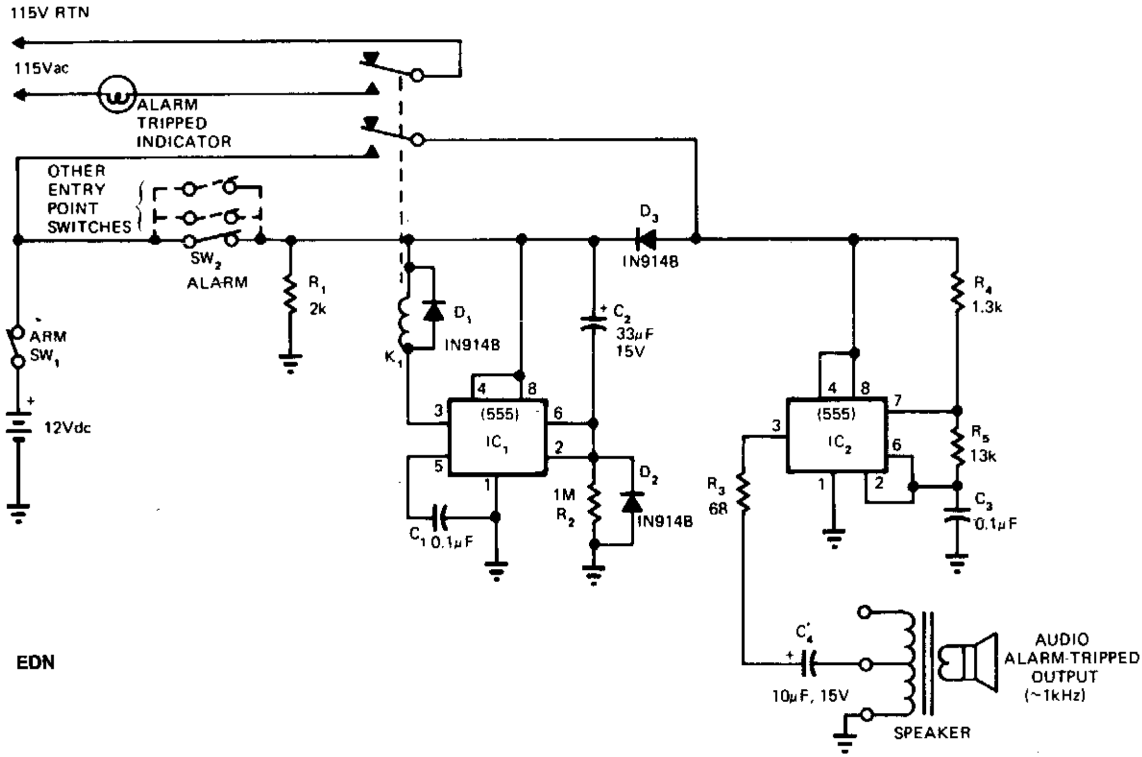
Fig. 2-5

The heart of the circuit is a 555 oscillator/timer, U1, configured for monostable operation. The output of U1 at pin 3 is tied to the gate of SCR1. As long as S1 – S5, which are connected to the trigger input of U1, are open, the circuit remains in the ready state, and does not trigger SCR1 into conduction. Because the relay is not energized, battery current is routed through the relay's normally-closed terminal and through current-limiting resistor R3 to LED2, causing it to light.

However, when one of the switches (S1 – S5) is closed, grounding U1 pin 2, the output of U1 at pin 3 increases, activating SCR1. That energizes the relay, pulling the wiper of K1 to the normally-open terminal, causing LED1 to light and BZ1 to sound.

The duration of the output is determined by the RC time-constant circuit, formed by R1 and C1. Resistor R2 regulates the output of U1 to a safe value for the gate of SCR1. Switches S1 – S5 are to doors, windows, etc. A switch can be connected in series with B1 to activate and deactivate the alarm circuit when it's not needed.

BURGLAR ALARM



EDN

Fig. 2-6

This circuit cannot be shut off for 10 to 60 seconds—even if the trip condition is immediately removed. It draws no standby power from the battery and is self-resetting.

3

Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

INPUT/OUTPUT BUFFER AMPLIFIER FOR ANALOG MULTIPLEXERS

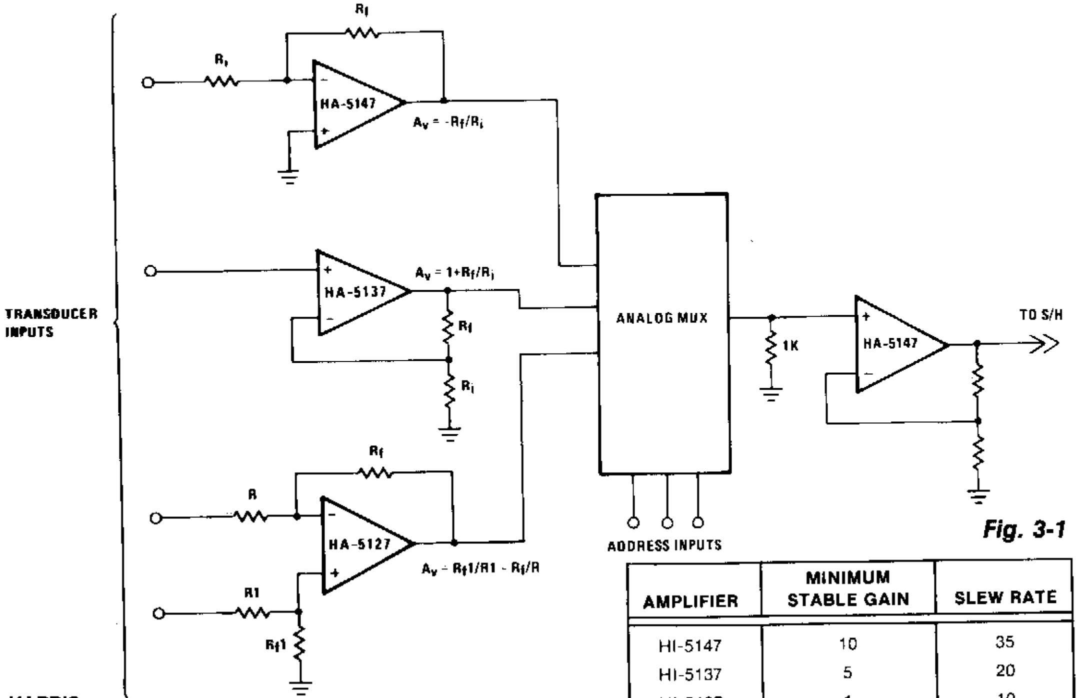


Fig. 3-1

AMPLIFIER	MINIMUM STABLE GAIN	SLEW RATE
HI-5147	10	35
HI-5137	5	20
HI-5127	1	10

The precision input characteristics of the HA-5147 help simplify system *error budgets*, while its speed and drive capabilities provide fast charging of the multiplexer's output capacitance. This speed eliminates an increased multiplexer acquisition time, which can be induced by more limited amplifiers. The HA-5147 accurately transfers information to the next stage while effectively reducing any loading effects on the multiplexer's output.

ABSOLUTE VALUE NORTON AMPLIFIER

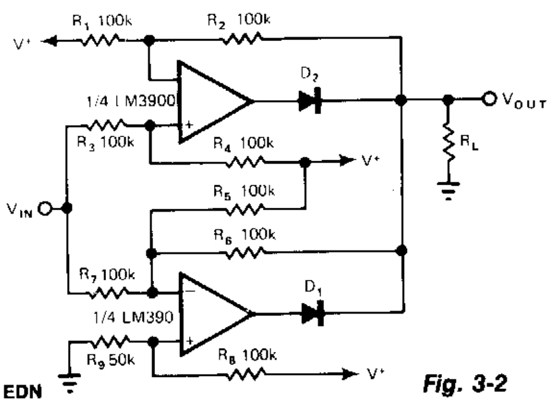
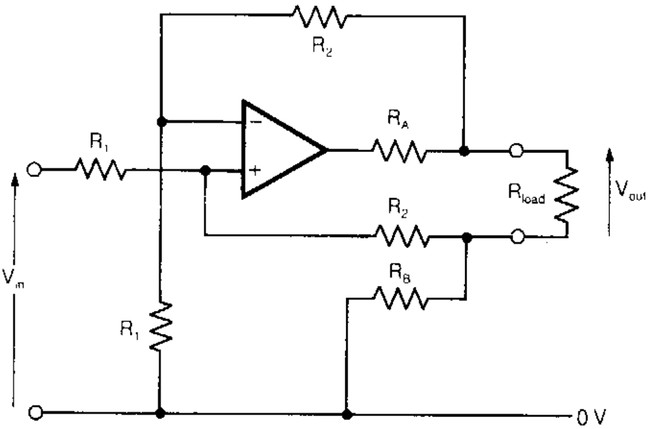


Fig. 3-2

The noninverting amplifier has a gain of R2/R3 (1 in this case) and produces a voltage of V_{out} during a positive excursion of V_{in} with respect to ground. The inverting amplifier accommodates the negative excursions of V_{in}; its gain is given by −R6/R7, which equals −1 to maintain symmetry with the noninverting amplifier. R9 provides adjustment for the symmetry, supply variations, and offsets. Even though the circuit operates on a single supply, V_{in} can go negative to the same extent that it goes positive.

INTRINSICALLY SAFE PROTECTED OP AMP



ELECTRONIC ENGINEERING

Fig. 3-3

In intrinsically safe applications, it is sometimes necessary to separate sections of circuitry by resistors which limit current under fault conditions. The circuit shown provides an accurate analogue output with effectively zero output impedance, despite having resistors in series with the output. The output voltage is given by:

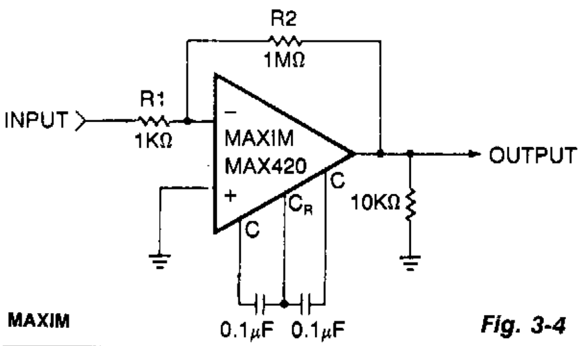
$$V_{out} = \frac{V_{in} R_2}{R_1}$$

which is independent of R_A and R_B. The values of R_A and R_B should be chosen to achieve the desired current limiting, but note that a proportion of the voltage given at the op-amp output will be dropped across these resistors. This limits the output swing at the load to approximately:

$$\frac{V_S R_{load}}{R_A + R_B + R_{load}}$$

where: V_S = voltage swing at the op-amp output. Any type of op amp would be suitable.

± 15-V CHOPPER AMPLIFIER



MAXIM

Fig. 3-4

This simple circuit is a gain-of-1000 inverting amplifier. It will amplify submillivolt signals up to signal levels suitable for further processing. In almost all system applications, it is best to use as much gain as possible in the MAX420, thus minimizing the effects of later-stage offsets. For example, if circuitry following the MAX420 has an offset of 5 mV, the additional offset referred back to the MAX420 input (gain = 1000) will be 5 μV, doubling the system's offset error.

COMPOSITE AMPLIFIER

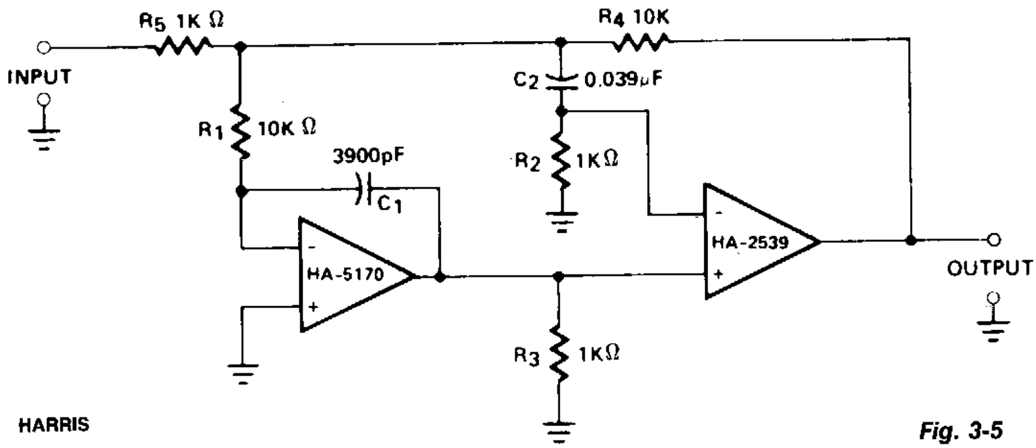


Fig. 3-5

A composite configuration greatly reduces dc errors without compromising the high-speed, wideband characteristics of HA-2539. The HA-2540 could also be used, but with slightly lower speeds and bandwidth response.

The HA-2539 amplifies signals above 40 kHz which are fed forward via C2; R2 and R5 set the voltage gain at −10. The slew rate of this circuit was measured at 350 V/μs. Settling time to a 0.1% level for a 10-V output step is under 150 ns and the gain bandwidth product is 300 MHz.

The HA-5170 amplifies signals below 40 kHz, as set by C1 and R1, and controls the dc input characteristics such as offset voltage, drift, and bias currents of the composite amplifier. Therefore, it has an offset voltage of 100 μV, drift of 2 μV/°C, and bias currents in the 20-pA range. The offset voltage can be externally nulled by connecting a 20-KΩ pot to pins 1 and 5, with the wiper tied to the negative supply. The dc gains of the HA-5170 and HA-2539 are cascaded; this means that the dc gain of the composite amplifier is well over 160 dB.

The excellent ac and dc performance of this composite amplifier is complemented by its low noise performance, 0.5-μV rms from 0.1 Hz to 100 Hz. It is very useful in high-speed data acquisition systems.

CASCADED AMPLIFIER

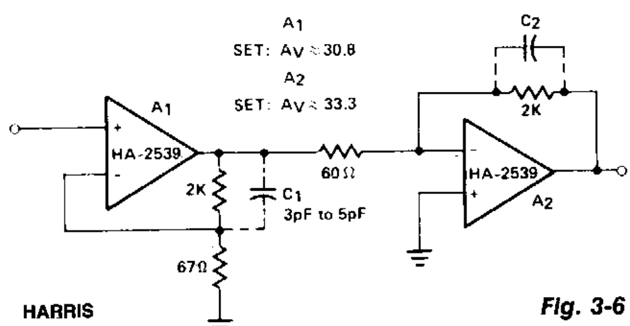


Fig. 3-6

Cascaded amplifier sections are used to extend bandwidth and increase gain. Using two HA-2539 devices, this circuit is capable of 60-dB gain at 20 MHz.

INVERTING AMPLIFIER

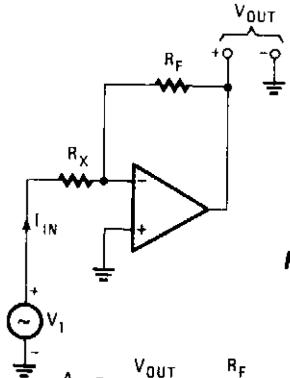


Fig. 3-7

$$A_V = -\frac{V_{OUT}}{V_1} = -\frac{R_F}{R_X}$$

HANDS-ON ELECTRONICS

NONINVERTING AMPLIFIER

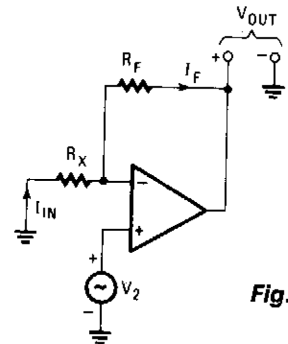
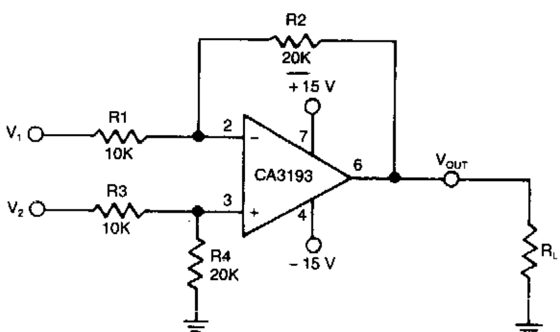


Fig. 3-8

$$A_V = \frac{V_{OUT}}{V_2} = \frac{R_F}{R_X} + 1$$

HANDS-ON ELECTRONICS

DIFFERENTIAL AMPLIFIER



ALL RESISTANCE VALUES ARE IN OHMS

$$V_{OUT} = V_2\left(\frac{R4}{R3 + R4}\right)\left(\frac{R1 + R2}{R1}\right) - V_1\left(\frac{R2}{R1}\right)$$

IF R4 = R2, R3 = R1 AND $\frac{R2}{R1} = \frac{R4}{R3}$

THEN $V_{OUT} = (V_2 - V_1)\left(\frac{R2}{R1}\right)$

FOR VALUES ABOVE $V_{OUT} = (V_2 - V_1)$

IF A_V IS TO BE MADE 1 AND IF R1 = R3 = R4 = R
WITH R2 = 0.999R (0.1% MISMATCH IN R2)

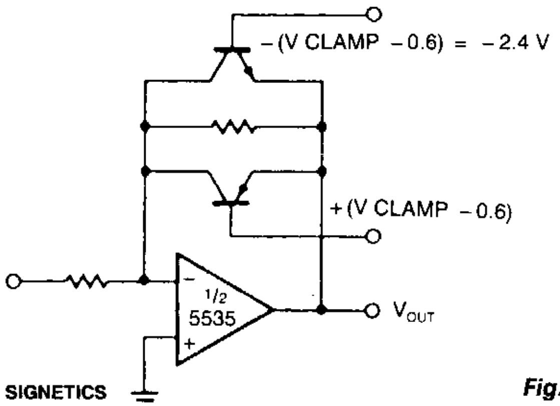
THEN $V_{OCM} = 0.0005\ V_{IN}$ OR CMRR = 66 dB
THUS, THE CMRR OF THIS CIRCUIT IS LIMITED BY THE MATCHING OR MISMATCHING OF THIS NETWORK RATHER THAN THE AMPLIFER

GE/RCA

Fig. 3-9

This differential amplifier uses a CA3193 BiMOS op amp. This classical, differential input-to-signal-ended output converter when used with low-resistance signal source will maintain level of CMRR, if $R_1 = R_3 + R_4$.

ACTIVE CLAMP-LIMITING AMPLIFIER

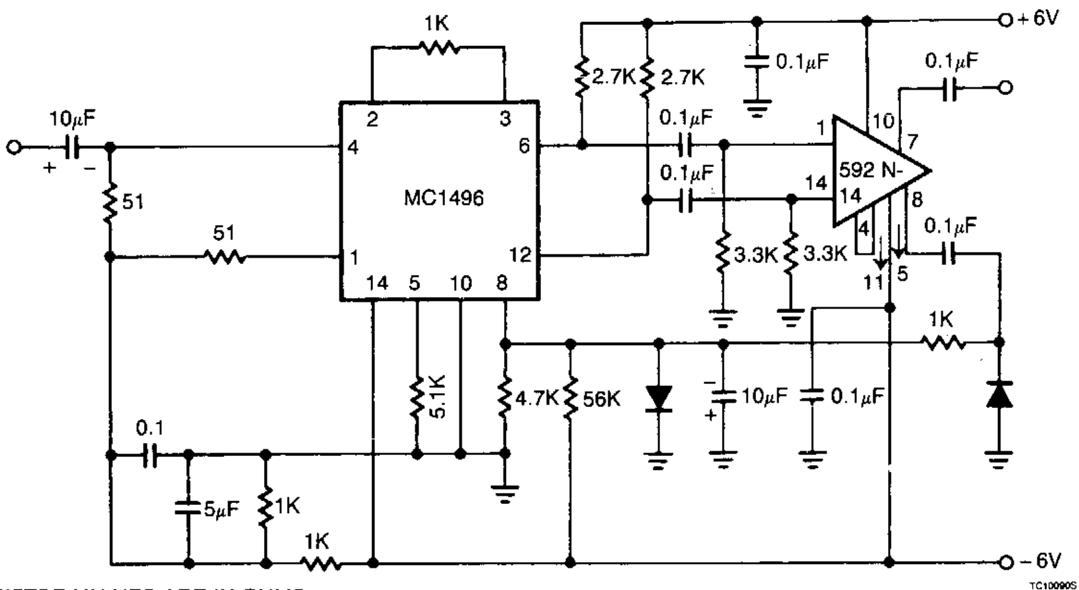


SIGNETICS

The modified inverting amplifier uses an active clamp to limit the output swing with precision. Allowance must be made for the V_{BE} of the transistors. The swing is limited by the base-emitter breakdown of the transistors. A simple circuit uses two back-to-back zener diodes across the feedback resistor, but tends to give less precise limiting and cannot be easily controlled.

Fig. 3-10

WIDE-BAND AGC AMPLIFIER



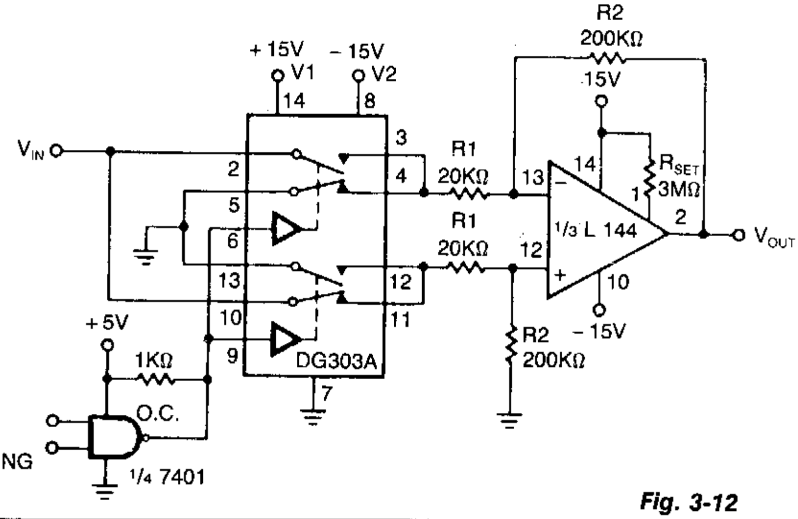
NOTE:
ALL RESISTOR VALUES ARE IN OHMS

SIGNETICS

Fig. 3-11

The NE592 is connected in conjunction with a MC1496 balanced modulator to form an excellent automatic gain control system. The signal is fed to the signal input of the MC1496 and rc-coupled to the NE592. Unbalancing the carrier input of the MC1496 causes the signal to pass through unattenuated. Rectifying and filtering one of the NE592 outputs produces a dc signal which is proportional to the ac signal amplitude. After filtering, this control signal is applied to the MC1496, causing its gain to change.

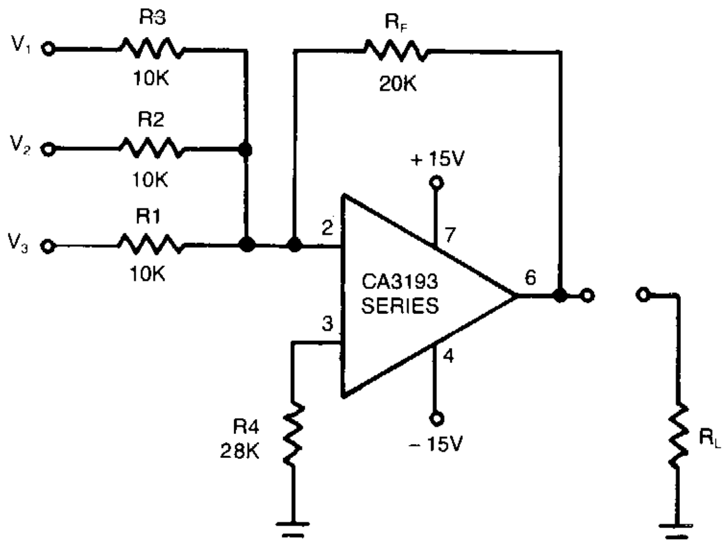
POLARITY-REVERSING LOW-POWER AMPLIFIER



SILICONIX

Fig. 3-12

SUMMING AMPLIFIER



$$V_{OUT} = -\left(\frac{R_F}{R1}V_1 + \frac{R_F}{R2}V_2 + \frac{R_F}{R3}V_3\right)$$

$$V_{OUT} = -(2V_1 + 2V_2 + 2V_3)$$

ALL RESISTANCE VALUES ARE IN OHMS

GE/RCA

Fig. 3-13

This circuit uses a CA3193 BiMOS op amp. Because input noise of the amplifier is increased by R_F/R1//R2//R3, and the gain that a single input will amplify is the gain of only one of the input channels (R_F/R1), for good noise performance, use the smallest number of inputs.

AC-COUPLED DYNAMIC AMPLIFIER

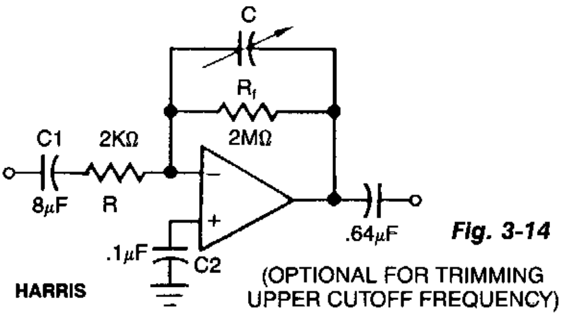


Fig. 3-14

This circuit acts as a bandpass filter with gain and would be most useful for biomedical instrumentation. Low-frequency cutoff is set at 10 Hz while the high-frequency breakpoint is given by the open-loop rolloff characteristic of the HA-5141/42/44. In this case, the A_{VCL} = 60 dB where the rolloff occurs at approximately 300 Hz. This corner frequency may be trimmed by inserting a capacitor in parallel with R_f.

FORWARD-CURRENT BOOSTER

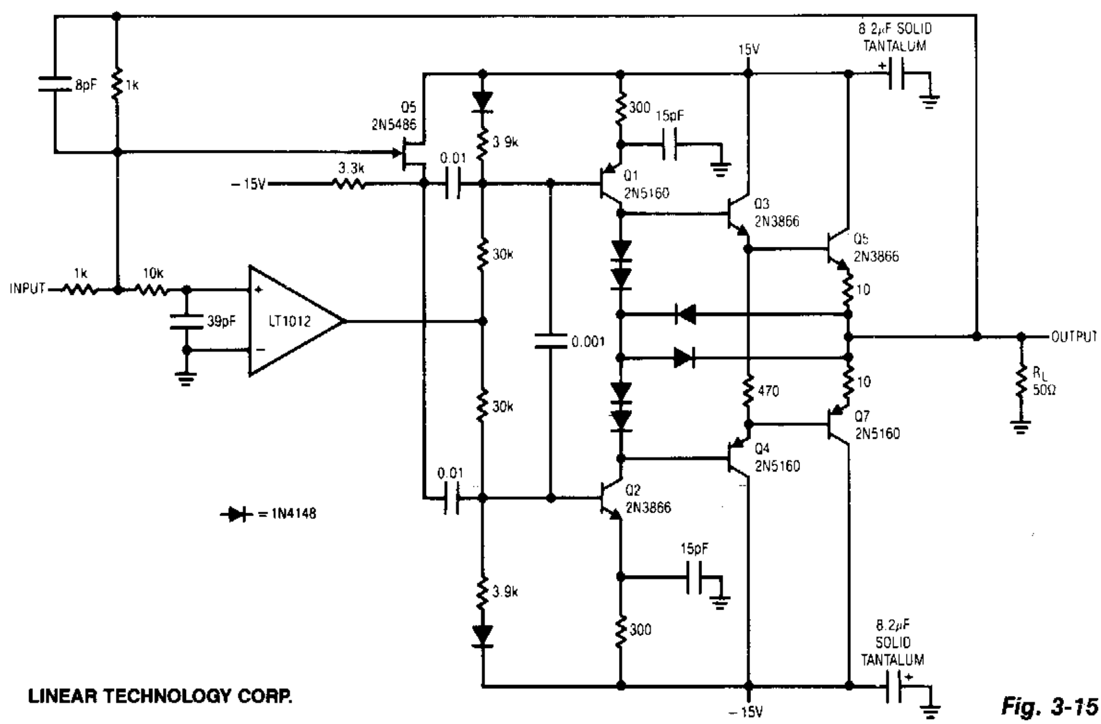


Fig. 3-15

The LT1012 corrects dc errors in the booster stage, and does not set high-frequency signals. Fast signals are fed directly to the stage via Q5 and the 0.01-µF coupling capacitors. Dc and low-frequency signals drive the stage via the op-amp's output. The output stage consists of current sources, Q1 and Q2, driving the Q3 – Q5 and Q4 – Q7 complementary emitter follows. The diode network at the output steers drive away from the transistor bases when output current exceeds 250 mA, providing fast short-circuit protection. The circuit's high frequency summing node is the junction of the 1-K and 10-K resistors at the LT1012. The 10 K/39 pF pair filters high frequencies, permitting accurate dc summation at the LT1012's positive input. This current-boosted amplifier has a slew rate in excess of 1000 V/µs, a full power bandwidth of 7.5 MHz and a 3-dB point of 14 MHz.

DC-STABILIZED FAST AMPLIFIER

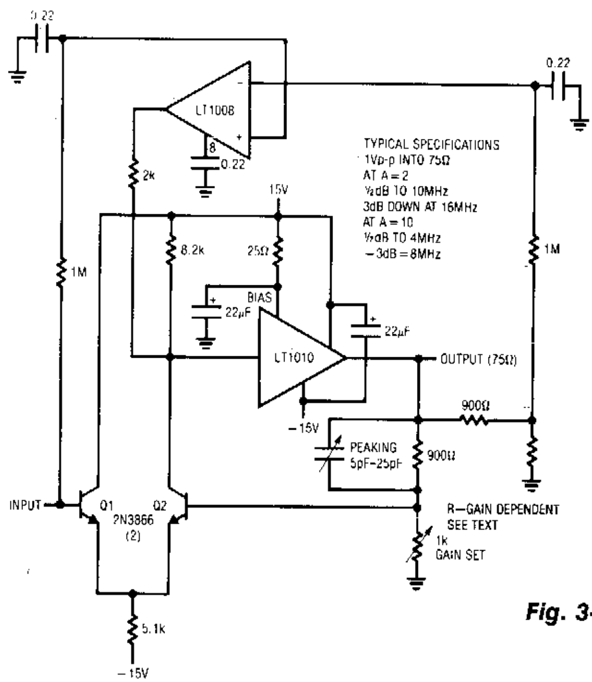
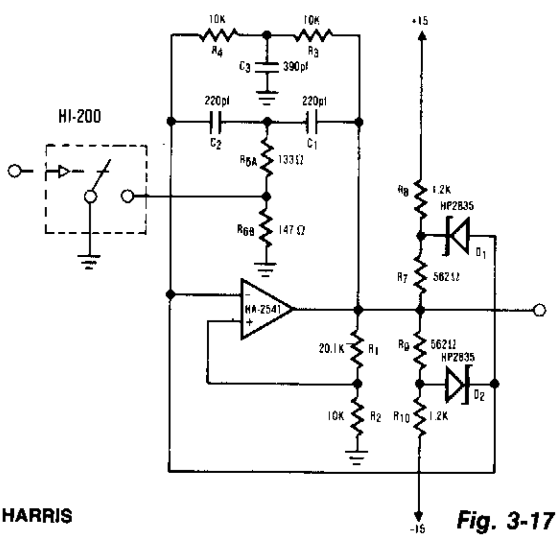


Fig. 3-16

LINEAR TECHNOLOGY CORP.

This amplifier functions over a wide range of gains, typically 1 – 10. It combines the LT1010 and a fast discrete stage with an LT1008 based dc stabilizing loop. Q1 and Q2 form a differential stage which single-ends into the LT1010. The circuit delivers 1 V pk-pk into a typical 75-Ω video load. At $A = 2$, the gain is within 0.5 dB to 10 MHz with the −3-dB point occurring at 16 MHz. At $A = 10$, the gain is flat (±0.5 dB to 4 MHz) with a −3-dB point at 8 MHz. The peaking adjustment should be optimized under loaded output conditions. This is a simple stage for fast applications where relatively low output swing is required. Its 1 V pk-pk output works nicely for video circuits. A possible problem is the relatively high bias current, typically 10 μA. Additional swing is possible, but more circuitry is needed.

WRITE AMPLIFIER

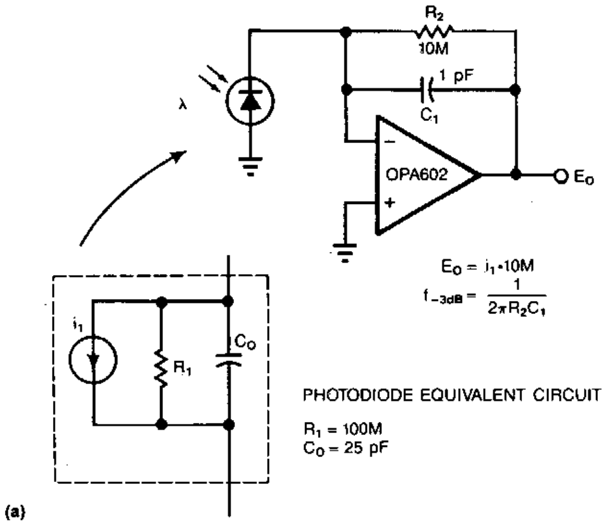


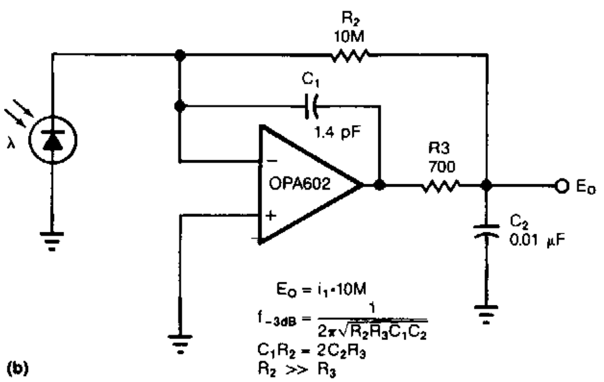
HARRIS

Fig. 3-17

The proliferation of industrial and computerized equipment containing programmable memory has increased the need for reliable recording media. The magnetic tape medium is presently one of the most widely used methods. The primary component of any magnetic recording mechanism is the write mechanism. The concept of the write generator is very basic. The digital input causes both a change in the output amplitude, as well as a change in frequency. This type of operation is accomplished by altering the value of a resistor in the standard twin-tee oscillator. A HI-201 analog switch was used to facilitate the switching action. The effect of the external components on the feedback network requires R6A and R6B to be much smaller than would normally have been expected when using the twin-tee feedback scheme.

LOW-NOISE PHOTODIODE AMPLIFIER





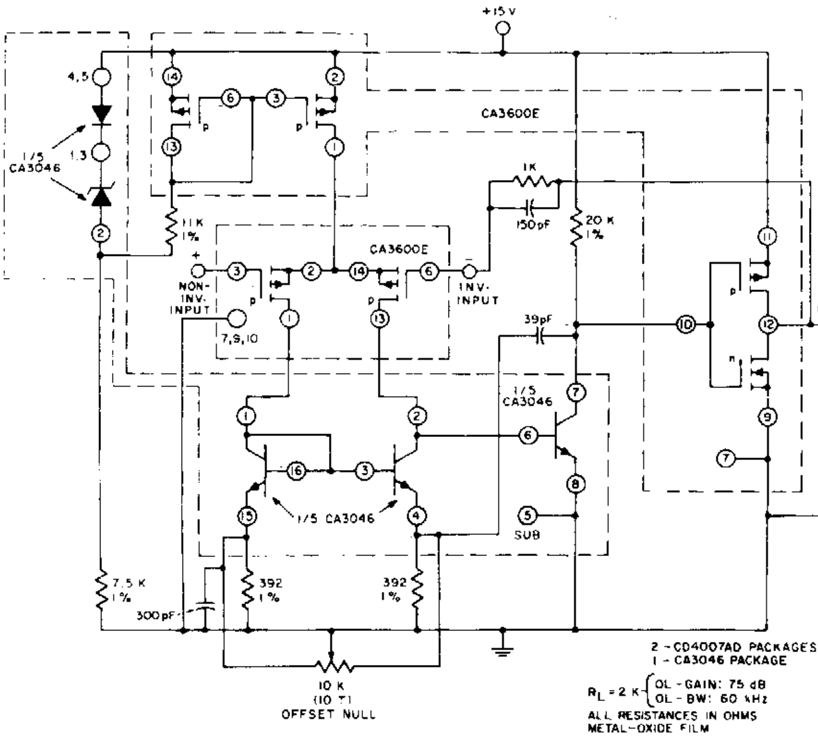
EDN *Fig. 3-18*

Adding two passive components to a standard photodiode amplifier reduces noise. Without the modification, the shunt capacitance of the photodiode reacting with the relatively large feedback resistor of the transimpedance (current-to-voltage) amplifier, creates excessive noise gain.

The improved circuit, Fig. 3-18b, adds a second pole, formed by R3 and C2. The modifications reduce noise by a factor of 3. Because the pole is within the feedback loop, the amplifier maintains its low output impedance. If you place the pole outside the feedback loop, you have to add an additional buffer, which would increase noise and dc error.

The signal bandwidth of both circuits is 16 kHz. In the standard circuit (Fig. 3-18a), the 1-pF stray capacitance in the feedback loop forms a single 16-kHz pole. The improved circuit has the same bandwidth as the first, but exhibits a 2-pole response.

VOLTAGE-FOLLOWER AMPLIFIER FOR SIGNAL-SUPPLY OPERATION



INTERSIL ***Fig. 3-19***

This unity-gain follower amplifier has a CMOS p-channel input, an npn second-gain stage, and a CMOS inverter output. The IC building blocks are two CA3600E's (CMOS transistor pairs) and a CA3046 npn transistor array. A zener-regulated leg provides bias for a 400-μA p-channel source, feeding the input stage, which is terminated in an npn current mirror. The amplifier voltage-offset is nulled with the 10-KΩ balance potentiometer. The second-stage current level is established by the 20-KΩ load, and is selected to approximately the first-stage current level, to assure similar positive and negative slew rates. The CMOS inverter portion forms the final output stage and is terminated in a 2-KΩ load, a typical value used with monolithic op amps. Voltage gain is affected by the choice of load resistance value. The output stage of this amplifier is easily driven to within 1 mV of the negative supply voltage.

CURRENT-SHUNT AMPLIFIER

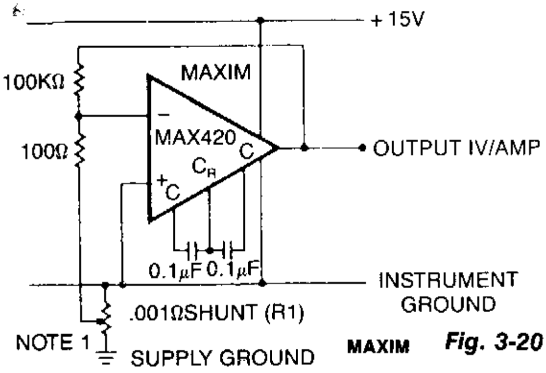
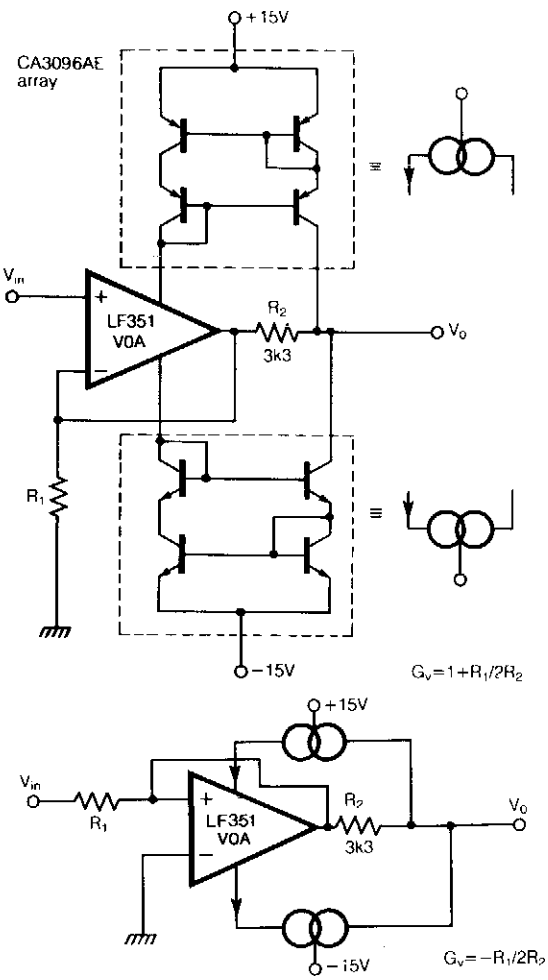


Fig. 3-20

This circuit measures the power-supply current of a circuit without really having a current-shunt resistor: R1 is only 3 cm of #20 gauge copper wire. A length of the power distribution wiring can be used for R1. The MAX420's CMVR includes its own negative power supply; therefore, it can both be powered by and measure current in the ground line.

CONSTANT-BANDWIDTH AMPLIFIER



The traditional restriction of constant gain-bandwidth products for a voltage amplifier can be overcome by employing feedback around a current amplifier. Two current mirrors, constructed from transistors in a CA3096AE array, effectively turn the LF351 op amp into a current amplifier. Feedback is then applied by using R2 and R1, turning the whole circuit into a feedback voltage amplifier with a noninverting gain of G of $1 + R_1/2R_2$.

Using the values shown, a constant bandwidth of 3.5 MHz is obtained for all voltage gains up to and beyond 100 at 10 V pk-pk output, equivalent to a gain-bandwidth product of 350 MHz from an op amp with an advertised unity gain-bandwidth of 10 MHz. An inverting gain configuration is also possible (see Fig. 2) where $G = R_1/2R_2$. Slewing rates are significantly improved by this approach; even a 741 can manage 100 V μs under these conditions since its output is a virtual earth. However, because the new configurations use current feedback to achieve bandwidth independence, an output buffer should be added for circuits where a significant output current is required.

Fig. 3-21

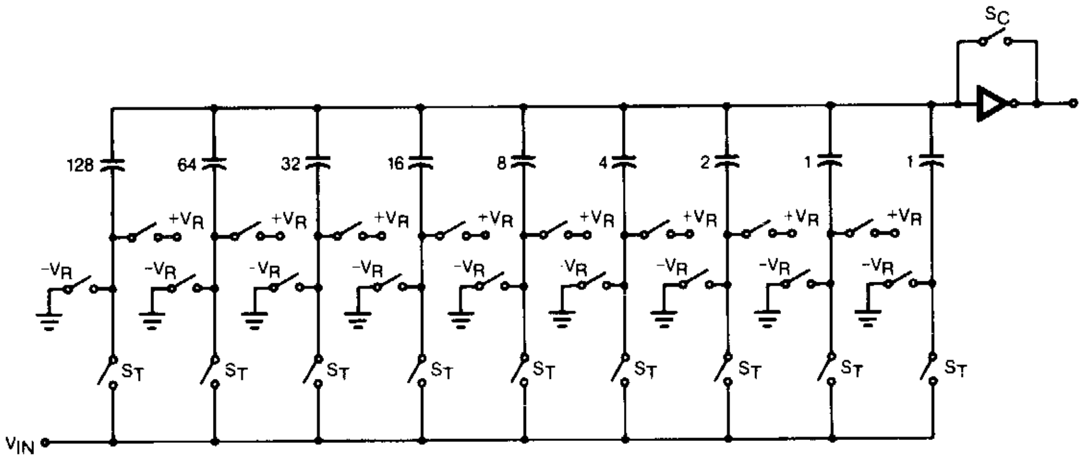
ELECTRONIC ENGINEERING

4

Analog-to-Digital Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SWITCHED-CAPACITOR ADC



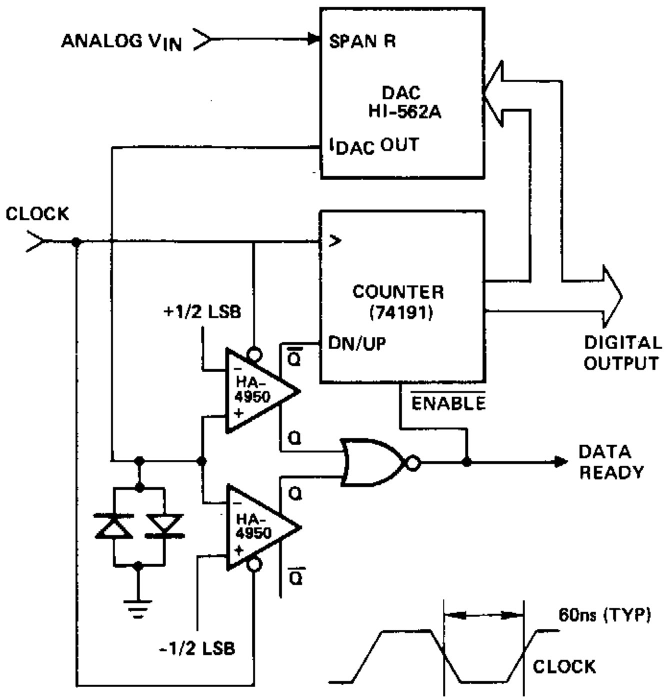
SILICONIX

Fig. 4-1

The CMOS comparator in the successive-approximation system determines each bit by examining the charge on a series of binary-weighted capacitors. In the first phase of the conversion process, the analog input is sampled by closing switch *SC* and all *ST* switches, and by simultaneously charging all the capacitors to the input voltage.

In the next phase of the conversion process, all *ST* and *SC* switches are opened and the comparator begins identifying bits by identifying the charge on each capacitor relative to the reference voltage. In the switching sequence, all 8 capacitors are examined separately until all 8 bits are identified, and then the charge-convert sequence is repeated. In the first step of the conversion phase, the comparator looks at the first capacitor (binary weight = 128). One pole of the capacitor is switched to the reference voltage, and the equivalent poles of all the other capacitors on the ladder are switched to ground. If the voltage at the summing node is greater than the trip point of the comparator—approximately 1/2 the reference voltage, a bit is placed in the output register, and the 128-weight capacitor is switched to ground. If the voltage at the summing node is less than the trip point of the comparator, this 128-weight capacitor remains connected to the reference input through the remainder of the capacitor-sampling (bit-counting) process. The process is repeated for the 64-weight capacitor, the 32-weight capacitor, and so forth down the line, until all bits are tested. With each step of the capacitor-sampling process, the initial charge is redistributed among the capacitors. The conversion process is successive-approximation, but relies on charge shifting rather than a successive-approximation register—and reference d/a—to count and weigh the bits from MSB to LSB.

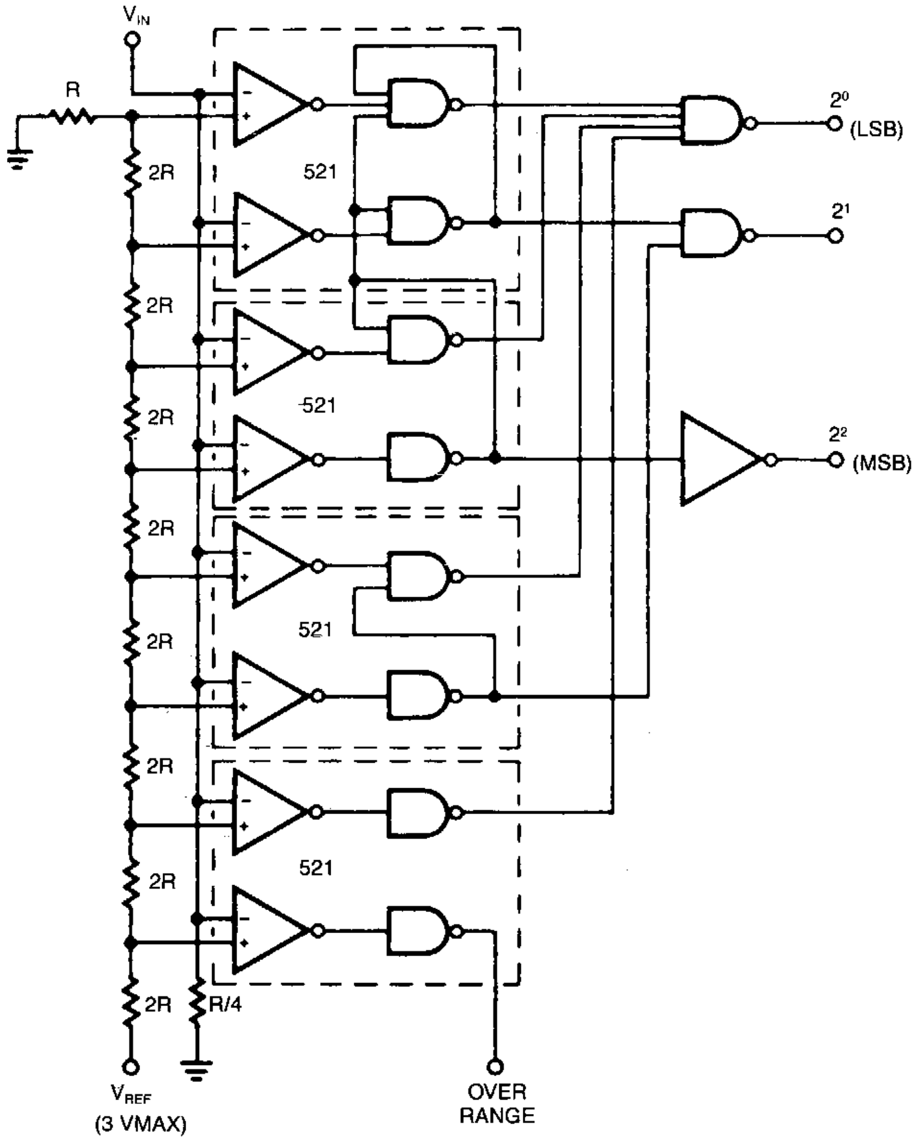
TRACKING ADC



HARRIS *Fig. 4-2*

The analog input is fed into the span resistor of a DAC. The analog input voltage range is selectable in the same way as the output voltage range of the DAC. The net current flow through the ladder termination resistance; i.e., 2 KΩ for HI-562A; produces an error voltage at the DAC output. This error voltage is compared with 1/2 LSB by a comparator. When the error voltage is within ±1/2 LSB range, the Q output of the comparators are both low, which stops the counter and gives a data ready signal to indicate that the digital output is correct. If the error exceeds the ±1/2 LSB range, the counter is enabled and driven in an up or down direction depending on the polarity of the error voltage.

The digital output changes state only when there is a significant change in the analog input. When monitoring a slowly varying input, it is necessary to read the digital output only after a change has taken place. The data ready signal could be used to trigger a flip-flop to indicate the condition and reset it after readout. The main disadvantage of the tracking ADC is the time required to initially acquire a signal; for a 12-bit ADC, it could be up to 4096 clock periods. The input signal usually must be filtered so that its rate of change does not exceed the tracking range of the ADC—1 LSB per clock period.

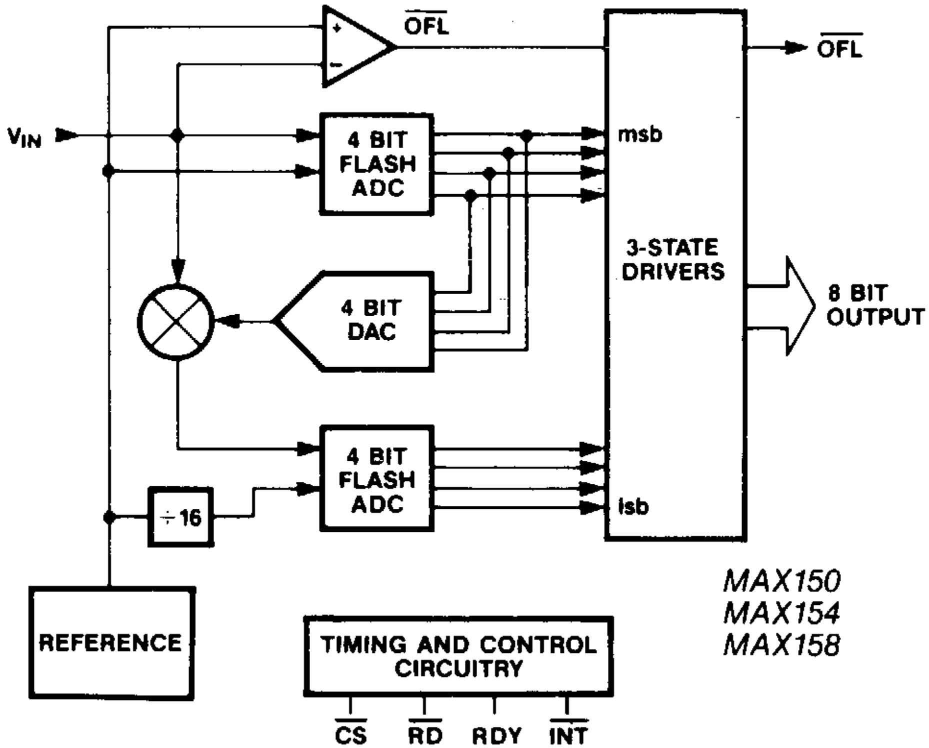


SIGNETICS

Fig. 4-3

Conversion speed of this design is the sum of the delay through the comparator and the decoding gates. Reference voltages for each bit are developed from a precision resistor ladder network. Values of R and $2R$ are chosen so that the threshold is 1/2 of the least significant bit. This assures maximum accuracy of ± 1/2 bit. The individual strobe line and duality features of the NE521 greatly reduced the cost and complexity of the design.

HALF-FLASH ADC



MAXIM

Fig. 4-4

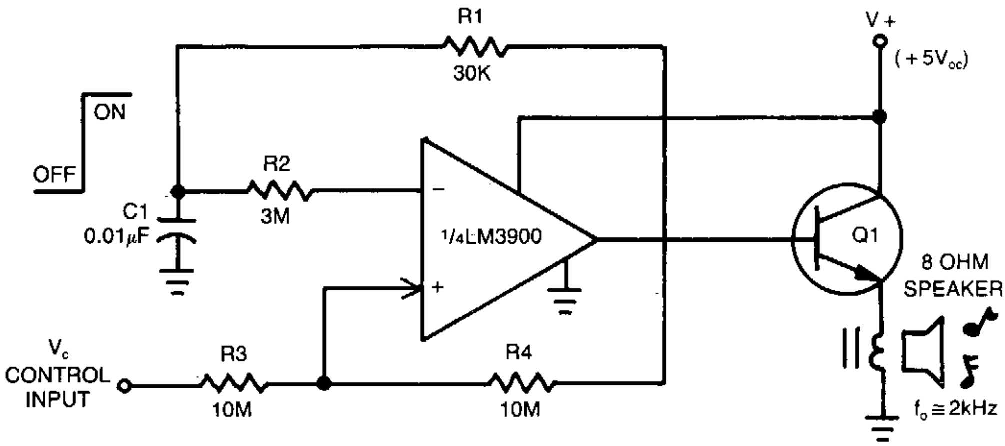
An a/d conversion technique which combines some of the speed advantages of flash conversion with the circuitry savings of successive approximation is termed *half-flash*. In an 8-bit, half-flash converter, two 4-bit flash a/d sections are combined. The upper flash a/d compares the input signal to the reference and generates the upper 4 data bits. This data goes to an internal DAC, whose output is subtracted from the analog input. Then, the difference can be measured by the second flash a/d, which provides the lower 4 data bits.

5

Annunciator

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

TRANSFORMERLESS TONE ANNUNCIATOR



EDN

Fig. 5-1

This circuit does not require an output transformer or an output coupling capacitor; the annunciator can easily be turned on or off by a control input voltage driving a 10-MΩ input resistor, R3. For a smaller acoustic output, replace output transistor, Q1, with a 100-Ω resistor, while also raising the voice coil impedance to 100 Ω, to prevent loading of the IC.

6

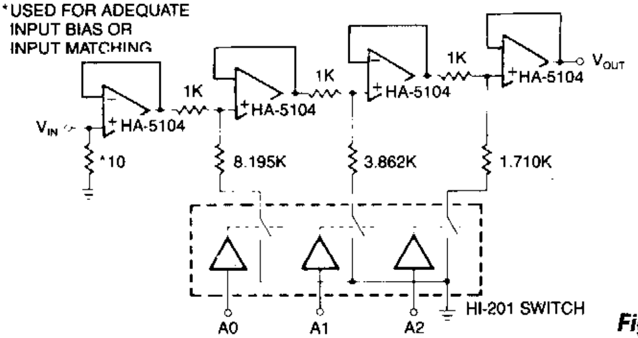
Attenuators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Digitally Programmable Attenuator
Programmable Attenuator
Voltage-Controlled Attenuator

DIGITALLY PROGRAMMABLE ATTENUATOR

A2	A1	A0	ATTENUATION
0	0	0	7dB
0	0	1	6dB
0	1	0	5dB
0	1	1	4dB
1	0	0	3dB
1	0	1	2dB
1	1	0	1dB
1	1	1	0dB

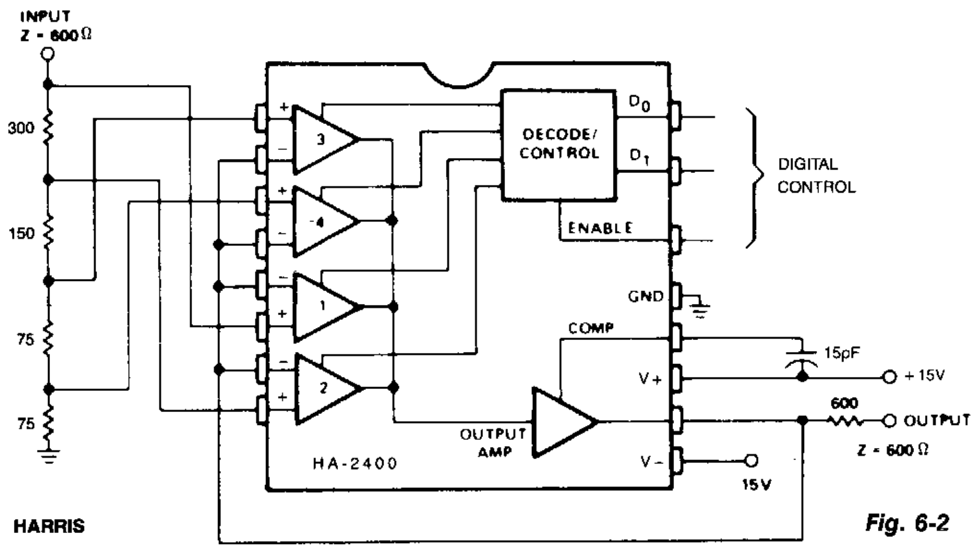


HARRIS

Fig. 6-1

The first stage is a simple buffer used to isolate the signal source from the attenuator stages to follow. Each of the subsequent stages is preceded by a voltage divider formed by two resistors and CMOS switch. Provided that the CMOS switch for each stage is closed, the drive signal will be attenuated according to the basic voltage divider relationship at each stage. In the event a switch is open, nearly all of the signal strength will be passed to the next stage through the 1-KΩ resistor. The amplifiers act as buffers for divider networks and reduce the interaction between stages. Eight levels of attenuation are possible with the circuit as illustrated, but more stages could be added. Each divider network must be closely matched to the resistor ratios shown or the level of attenuation will not match the levels in the logic chart.

PROGRAMMABLE ATTENUATOR

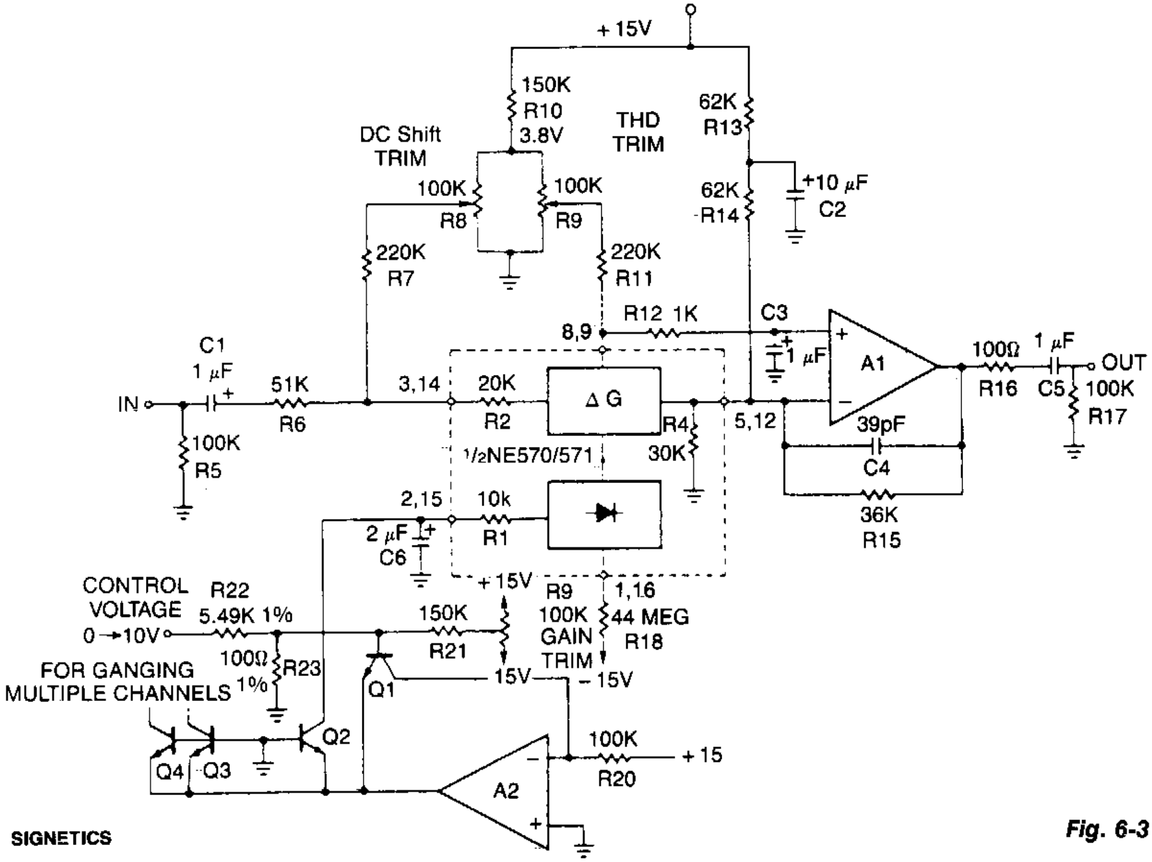


HARRIS

Fig. 6-2

This circuit performs the function of dividing the input signal by a selected constant (1, 2, 4, 8, etc.). While T, Z, or L sections could be used in the input attenuator, this is not necessary since the amplifier loading is negligible and a constant input impedance is maintained. The circuit is thus much simpler and more accurate than the usual method of constructing a constant impedance ladder, and switching sections in and out with analog switches. Two identical circuits can be used to attenuate a balanced line.

VOLTAGE-CONTROLLED ATTENUATOR



SIGNETICS

Fig. 6-3

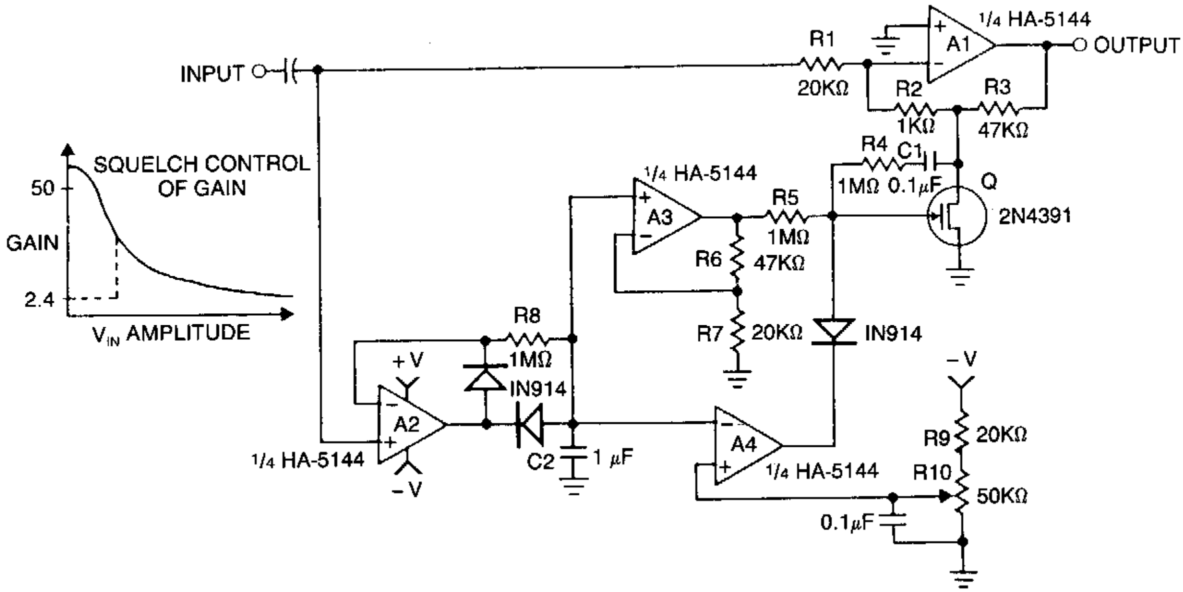
This typical circuit uses an external op amp for better performance and an exponential converter to get a control characteristic of −6 dB/V. Trim networks are shown to null out distortion and dc shift, and to fine trim the gain to 0 dB with 0 V of control voltage.

7

Audio Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

AGC WITH SQUELCH CONTROL

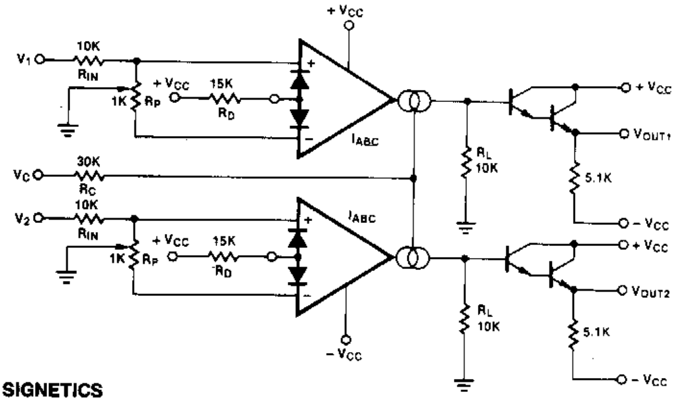


HARRIS

Fig. 7-1

Automatic gain control is a very useful feature in a number of audio amplifier circuits: tape recorders, telephone speaker phones, communication systems and PA systems. This circuit consists of a HA-5144 quad op amp and a FET transistor used as a voltage-controlled resistor to implement an AGC circuit with squelch control. The squelch function helps eliminate noise in communications systems when no signal is present and allows remote hands-free operation of tape recorder systems. Amplifier A1 is placed in an inverting-gain *T* configuration in order to provide a fairly wide gain range and a small signal level across the FET. The small signal level and the addition of resistors R5 and R6 help reduce nonlinearities and distortion. Amplifier A2 acts as a negative peak detector to keep track of signal amplitude. Amplifier A3 can be used to amplify this peak signal if the cutoff voltage of the FET is higher than desired. Amplifier A4 acts as a comparator in the squelch control section of the circuit. When the signal level falls below the voltage set by R10, the gate of the FET is pulled low—turning it off completely—and reducing the gain to 2.4. The output A4 can also be used as a control signal in applications, such as a hands-free tape recorder system.

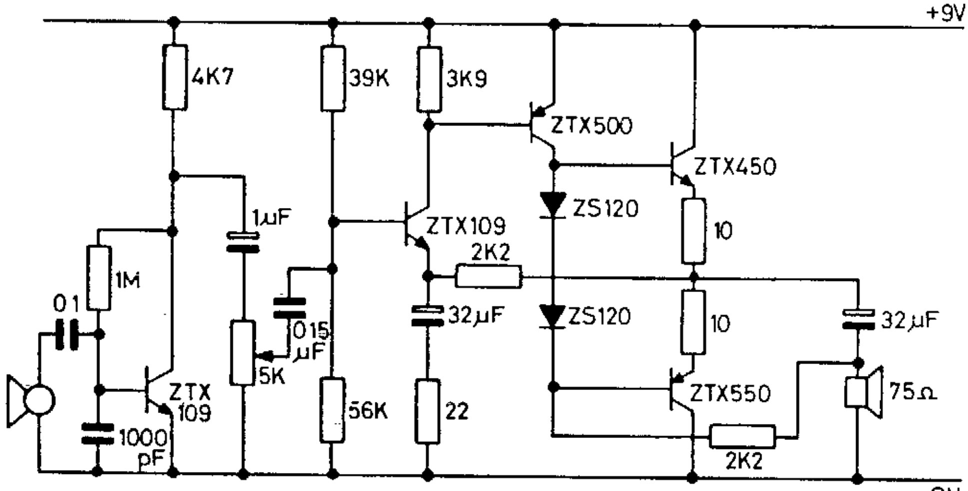
GAIN-CONTROLLED STEREO AMPLIFIER



SIGNETICS

Fig. 7-2

MICROPHONE AMPLIFIER



MICROPHONE 200Ω 60mH

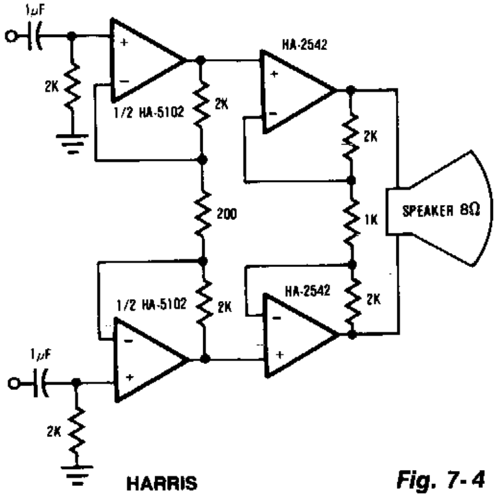
ZETEX, formerly FERRANTI

Fig. 7-3

This circuit features the ZTX450/ZTX550 transistors in a push-pull output stage. The following readings were taken at maximum volume:

Input: 0.4 mV rms
Output: 1.8 V rms
Voltage gain: 4500
Max. output before distortion: 2.25 V rms – supply current = 3.5 mA
Zero output-supply current: 3.5 mA
Wattage: 0.034 W
Frequency response: 250 Hz to 28 kHz

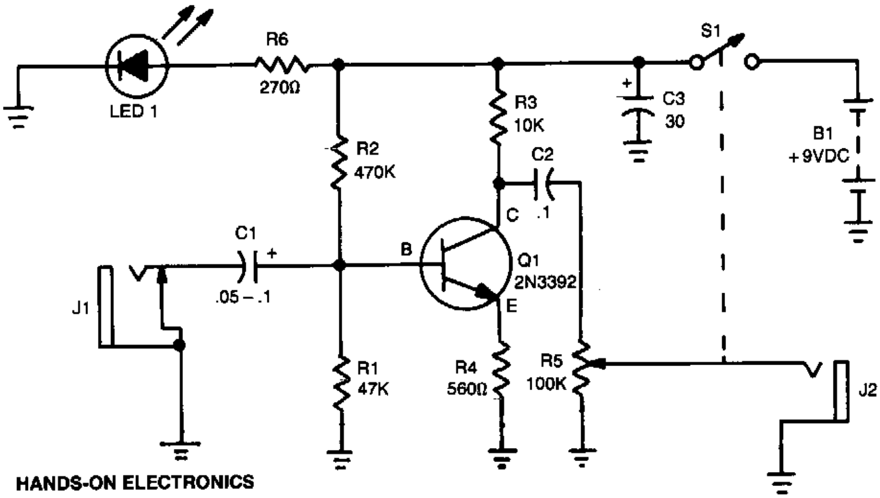
AUDIO CIRCUIT BRIDGE LOAD DRIVE



This circuit shows a method which increases the power capability of a drive system for audio speakers. Two HA-2542s are used to operate on half cycles only, which greatly increases their power handling capability. Bridging the speaker, as shown, makes 200 mA of output current available to drive the load. The HA-5102 is used as an ac-coupled, low noise preamplifier, which drives the bridge circuit.

HARRIS *Fig. 7-4*

20-dB AUDIO BOOSTER



HANDS-ON ELECTRONICS *Fig. 7-5*

The amplifier's gain is nominally 20 dB. Its frequency response is determined primarily by the value of just a few components—primarily C1 and R1. The values in the schematic diagram provide a response of ±3.0 dB from about 120 to over 20,000 Hz. Actually, the frequency response is flat from about 170 to well over 20,000 Hz; it's the low end that deviates from a flat frequency response. The low end's rolloff is primarily a function of capacitor C1, since R1's resistive value is fixed. If C1's value is changed to 0.1 μF, the low end's corner frequency—the frequency at which the low end rolloff starts—is reduced to about 70 Hz. If you need an even deeper low end rolloff, change C1 to a 1.0-μF capacitor. If it's an electrolytic type, make certain that it's installed into the circuit with the correct polarity – with the positive terminal connected to Q1's base terminal.

MICRO-SIZED AMPLIFIER

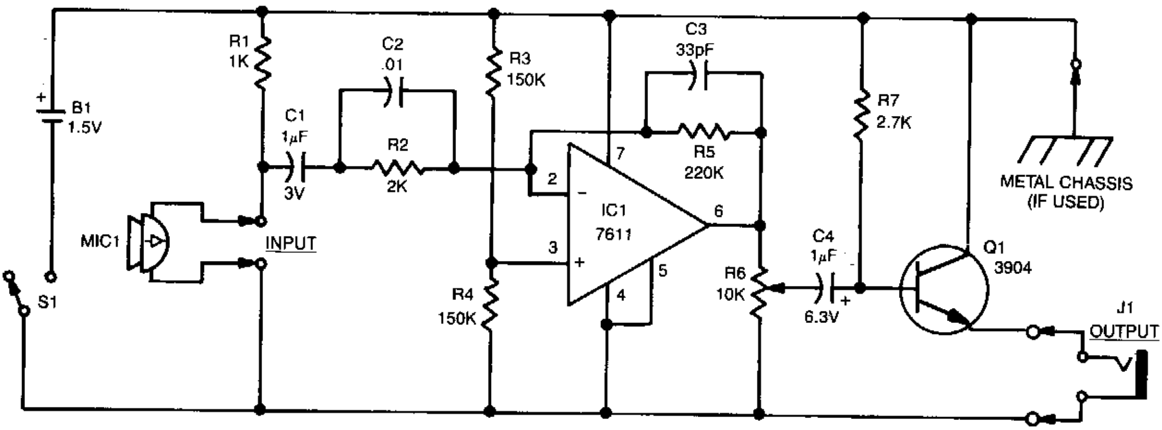
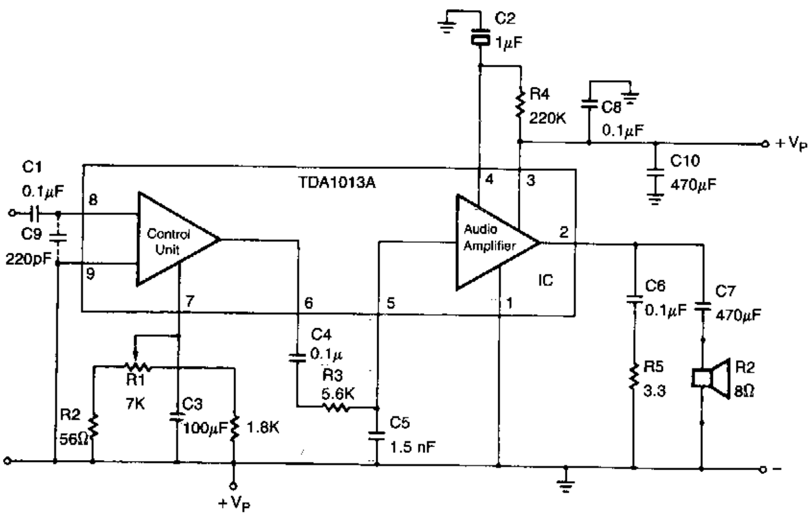


Fig. 7-6

Sound detected by electret microphone MIC1 is fed to IC1's input through resistor R2, and capacitors C1 and C2. Resistors R2 and R5 determine the overall stage gain, while C2 partially determines the amplifier's frequency response. To ensure proper operation, use a single-ended power supply. R3 and R4 simulate a null condition equal to half the power supply's voltage at IC1's noninverting input. The output of IC1 is transferred to emitter-follower amplifier Q1 via volume control R6. The high-Z-in/low-Z-out characteristic of the emitter-follower matches the moderately high-impedance output of IC1 to a low-impedance headphone load.

AUDIO AMPLIFIER

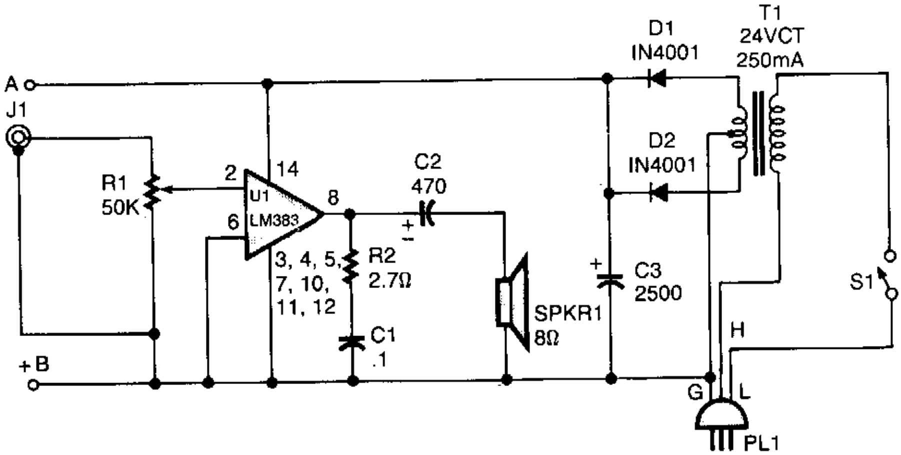


SIGNETICS

Fig. 7-7

C9 is necessary to filter-out rf input interferences. R3 in combination with C5 is used to limit the af frequency bandwidth. The 470-μF power supply decoupling capacitor is C10.

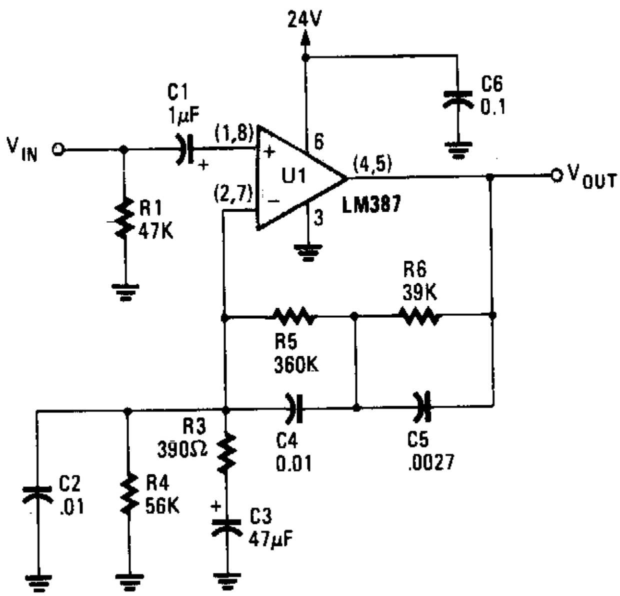
LINE-OPERATED AMPLIFIER



POPULAR ELECTRONICS *Fig. 7-8*

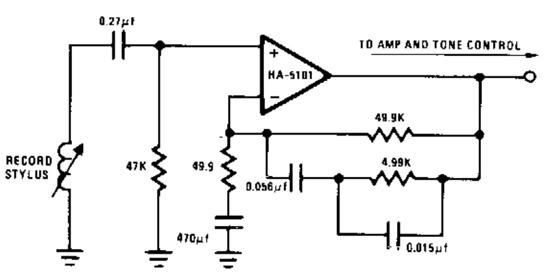
T1 isolates the unit from the line, and has a 24-V, center-tapped secondary. The output of the transformer is rectified by diodes D1 and D2 and filtered by capacitor C3 to provide 15 to 18 Vdc. The LM383 has built-in protection against speaker shorts.

MAGNETIC PHONO PREAMPLIFIER



POPULAR ELECTRONICS *Fig. 7-9*

RIAA PREAMPLIFIER

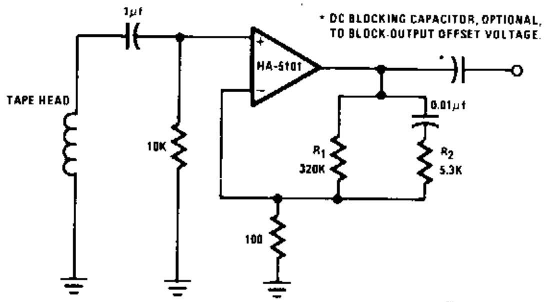


HARRIS

Fig. 7-10

The circuit essentially provides low-frequency boost below 318 Hz and high-frequency attenuation above 3150 Hz. Recent modifications to the response standard include a 31.5-Hz peak gain region to reduce dc-oriented distortion from external vibration.

PROFESSIONAL AUDIO NAB TAPE PLAYBACK PREAMPLIFIER

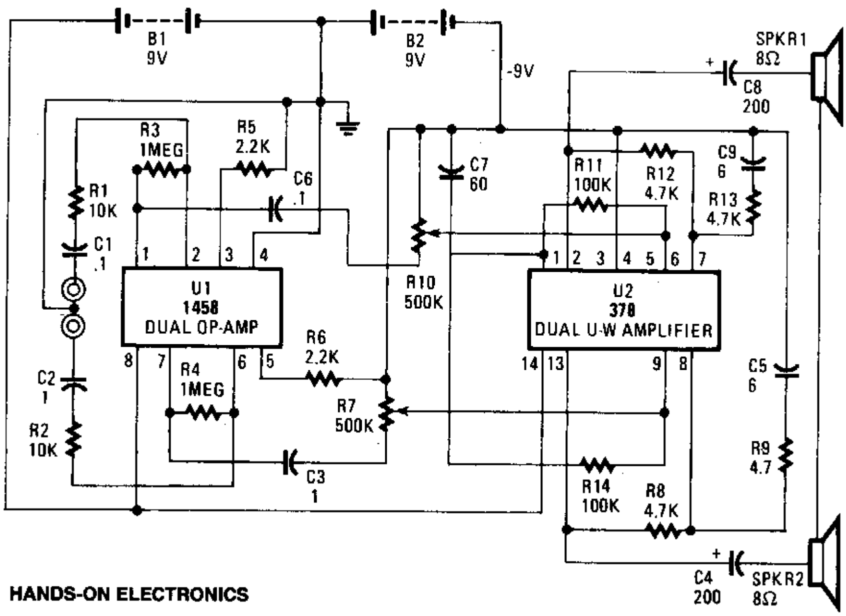


HARRIS

Fig. 7-11

The preamplifier is configured to provide low-frequency boost to 50 Hz, flat response to 3 kHz, and high-frequency attenuation above 3 kHz. Compensation for variations in tape and tape head performance can be achieved by trimming R1 and R2.

MINI-STEREO

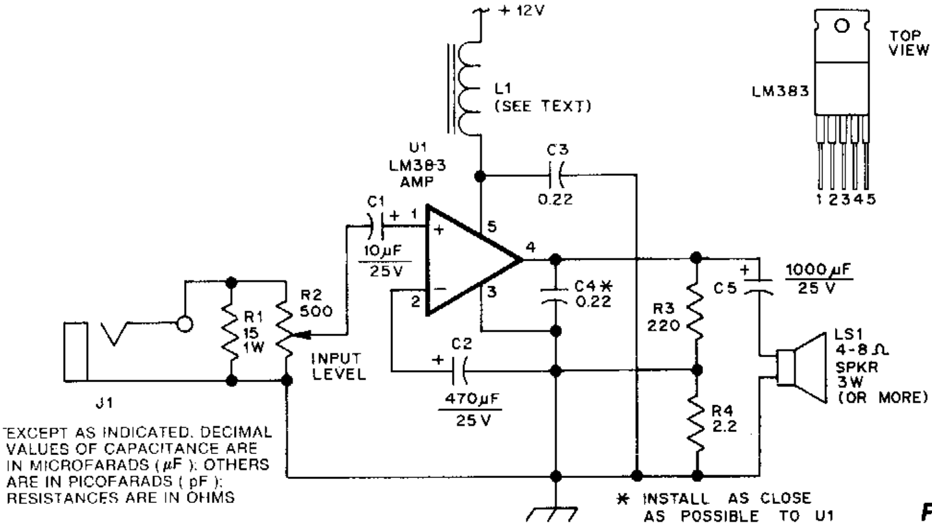


HANDS-ON ELECTRONICS

Fig. 7-12

This circuit is built around two chips: the MC1458 dual op amp, configured as a preamplifier, and the LM378 dual 4-watt amplifier. The gain of the preamp is given by R3/R1 for one side and R4/R2 for the other side, which is about 100. That gain can be varied by increasing the ratios. The left and right channel inputs are applied to pins 2 and 6. The left and right outputs of U1 at pins 7 and 2 are coupled through C5/R10 and C3/R6, respectively, to U2 to drive the two 8-Ω loudspeakers.

SPEAKER AMPLIFIER FOR HAND-HELD TRANSCEIVERS

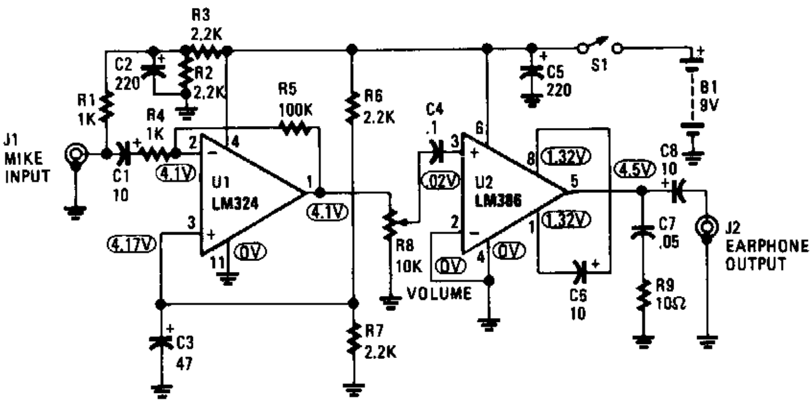


QST

Fig. 7-13

The LM383 is an audio-power amplifier that is capable of producing up to 8 W of audio output. R1 is essentially a load resistor for the hand-held transceiver's audio output. R2 can be composed of two fixed resistors in a 10:1 divider arrangement, but using a potentiometer makes it easy to set the amplifier's maximum gain. When powered from a vehicle's electrical system, the amplifier's +12V power source requires filter L1 to eliminate alternator whine. The LM383 can be mounted directly on the heatsink because the mounting tab is at ground potential.

TV AUDIO AMPLIFIER



HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 7-14

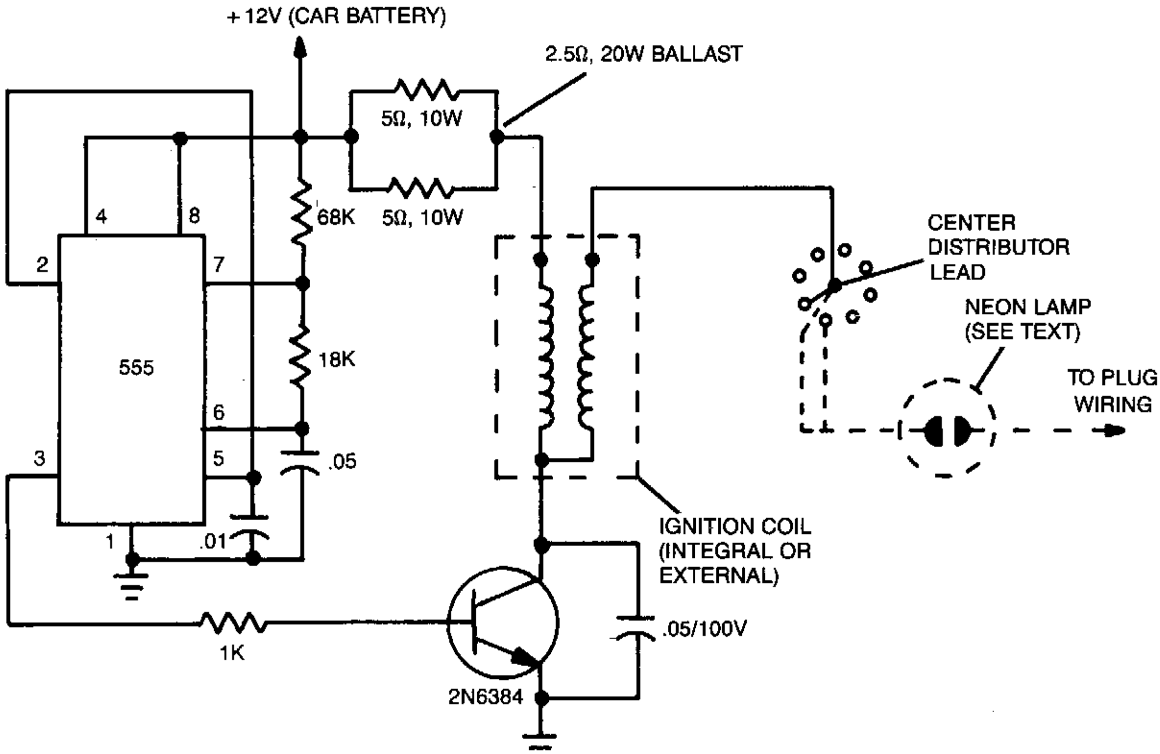
The amplifier picks up the TV's audio output signal and amplifies it to drive a set of earphones for private listening. It is built around an LM324 quad op amp and an LM386 low-power audio amplifier. The circuit uses an inexpensive electret microphone element as the pick-up and a set of earphones as the output device.

8

Automotive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

AUTOMOBILE IGNITION SUBSTITUTE



GERNSBACK PUBLICATIONS INC.

Fig. 8-1

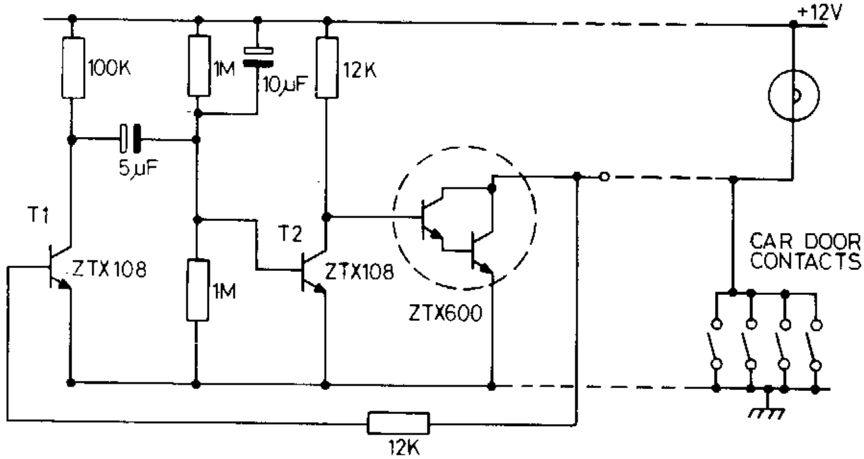
The ignition substitute provides a constant power source for the ignition coil. Its frequency, 0.5 – 1.0 kHz, is that used by an 8-cylinder engine with an idling speed of 650 RPM, and the unit provides a rapid spark at a 17% duty cycle, while staying within the power dissipation limits of the components.

The circuit consists of a 555 timer IC configured as an astable free-running multivibrator that is used to drive a high-current npn transistor, such as a 2N6384. The transistor should be heavily heatsinked because it might be drawing several amps over quite a long period of time.

The coil ballast can be from 0.68 to 6.5 Ω, depending on what's available. The 2.5-Ω, 20-W ballast shown works well. All the other resistors can be either 1/4- or 1/2-W devices, and the capacitor, between pins 1 and 5 of the 555, can range from 0.01 to 0.05 μF. Do not omit the 100-V, 0.05-μF capacitor across the transistor; it prevents voltage spikes from damaging the device.

Although designed for an 8-cylinder engine, this device can be used with other types. In addition, a neon bulb can be added to the circuit to verify the presence of a spark.

COURTESY LIGHT DELAY SWITCH

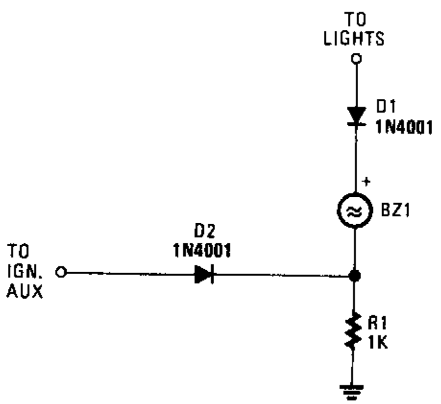


ZeTeX, formerly FERRANTI

Fig. 8-2

This circuit holds on the internal light for approximately one minute after the car doors are closed. When the door contacts open, a + VE pulse is applied to the base of T1. This transistor turns on, turning off T2 and charging the 10-μF capacitor. T3 turns on, holding on the internal light. The capacitor takes one minute to discharge when the circuit reverts to its original state.

LIGHTS-ON REMINDER



POPULAR ELECTRONICS

Fig. 8-3

With both the ignition and the car lights on, piezo transducer BZ1 draws no current and remains silent. With only the ignition on, diode D1 is reverse-biased and so prevents current flow through BZ1.

However, when the lights are on and the ignition is off, the transducer becomes energized and sounds to alert you to turn the lights off. With the ignition off and the lights on, D2 is reverse-biased, preventing current from flowing to the ignition. Resistor R1 prevents a short circuit when the ignition is on.

AUTOMOBILE LOCATER

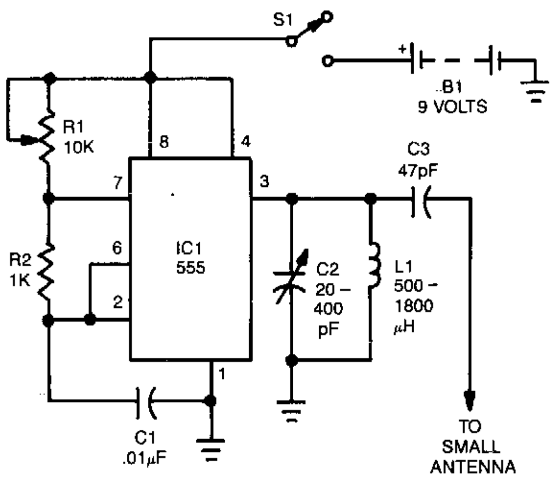


Fig. 8-4a

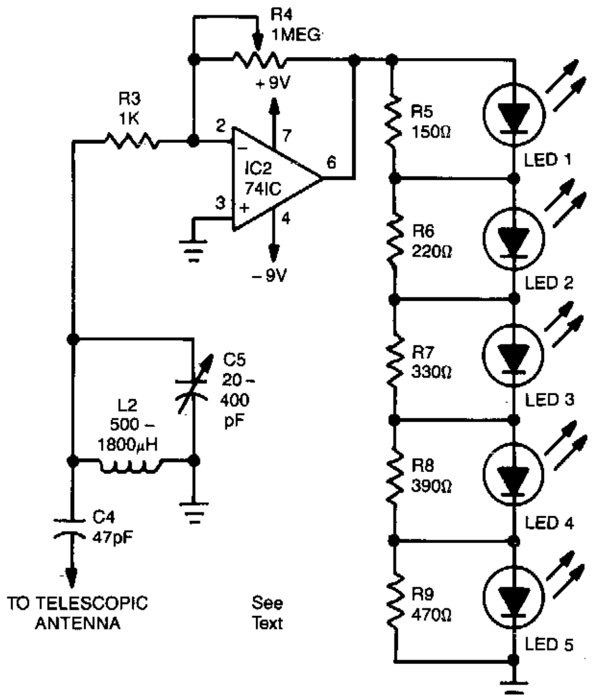


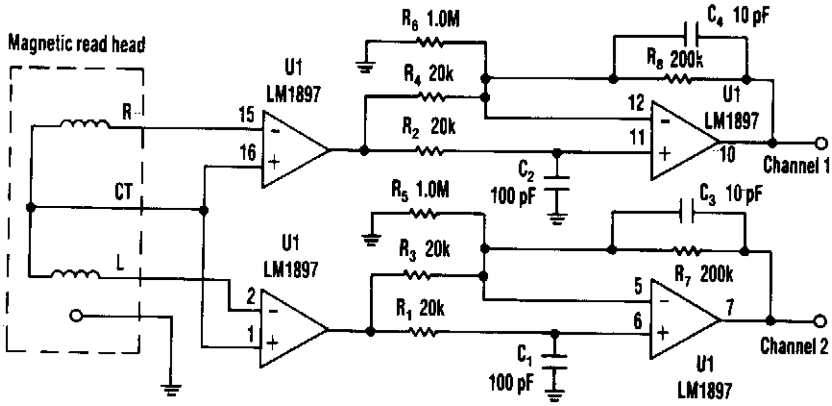
Fig. 8-4b

GERNSBACK PUBLICATIONS INC.

This locater is made up of two parts. The first is an rf oscillator, whose circuit is shown in Fig. 8-4a. The second is a sensitive receiver shown in Fig. 8-4b. The heart of the oscillator is a 555 timer IC. Tank circuit C2 and L1 is used to tune the transmitter. The antenna is coupled to the transmitter through C3. A telescopic antenna or a length of hookup wire will work quite well. At the receiver, the incoming signal is tuned by C5 and L2 before being passed on to the 741 IC. The five LEDs are used to indicate signal strength, they light up in order (1 to 5) as the signal gets stronger.

After the devices are built, the receiver and transmitter will need to be tuned. Tune the transmitter until all of the receiver's LEDs light. Separate the receiver and the transmitter—the farther apart they are the better—and adjust R4 until you get a maximum strength reading only when the receiver's antenna is pointed directly at the transmitter. Place the transmitter on the dashboard and completely extend the antenna. To find your car, just extend the telescope antenna to its full length and hold it parallel to the ground. Point the antenna to your far left, then swing it to your far right. Do that until you find in which direction the strongest signal lies, as indicated by the LEDs. The antenna will be pointing at your car.

READ-HEAD PREAMPLIFIER

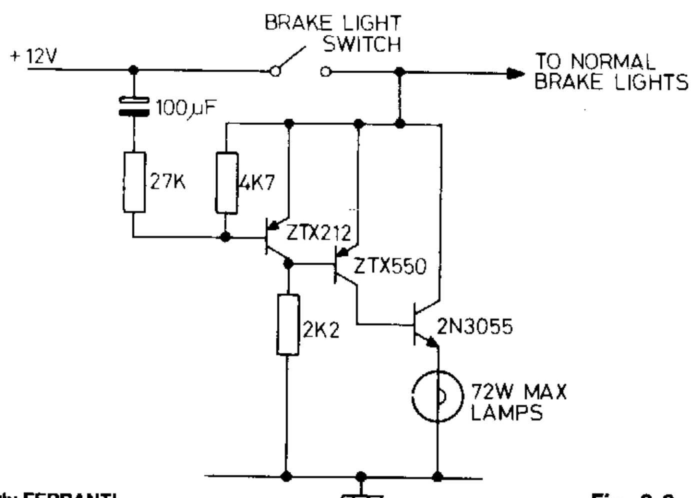


ELECTRONIC DESIGN

Fig. 8-5

Choosing dc rather than ac coupling can reduce much of the noise associated with preamplifiers for a magnetic reading head, particularly in the low frequencies. The LM1897 eliminates the need for the capacitor that usually ac couples the read head to the preamplifier input. The read head itself has a small resistance, typically 50 Ω, and so is less prone to noise pickup. Moreover, the LM1897 has a low-bias current; merely 2 μA as a worst case. Such a low-bias current flowing through the head's low resistance generates very little noise. Accordingly, even with a gain of 25, the first stage of the preamplifier circuit produces little noise.

DELAYED EXTRA BRAKE LIGHT

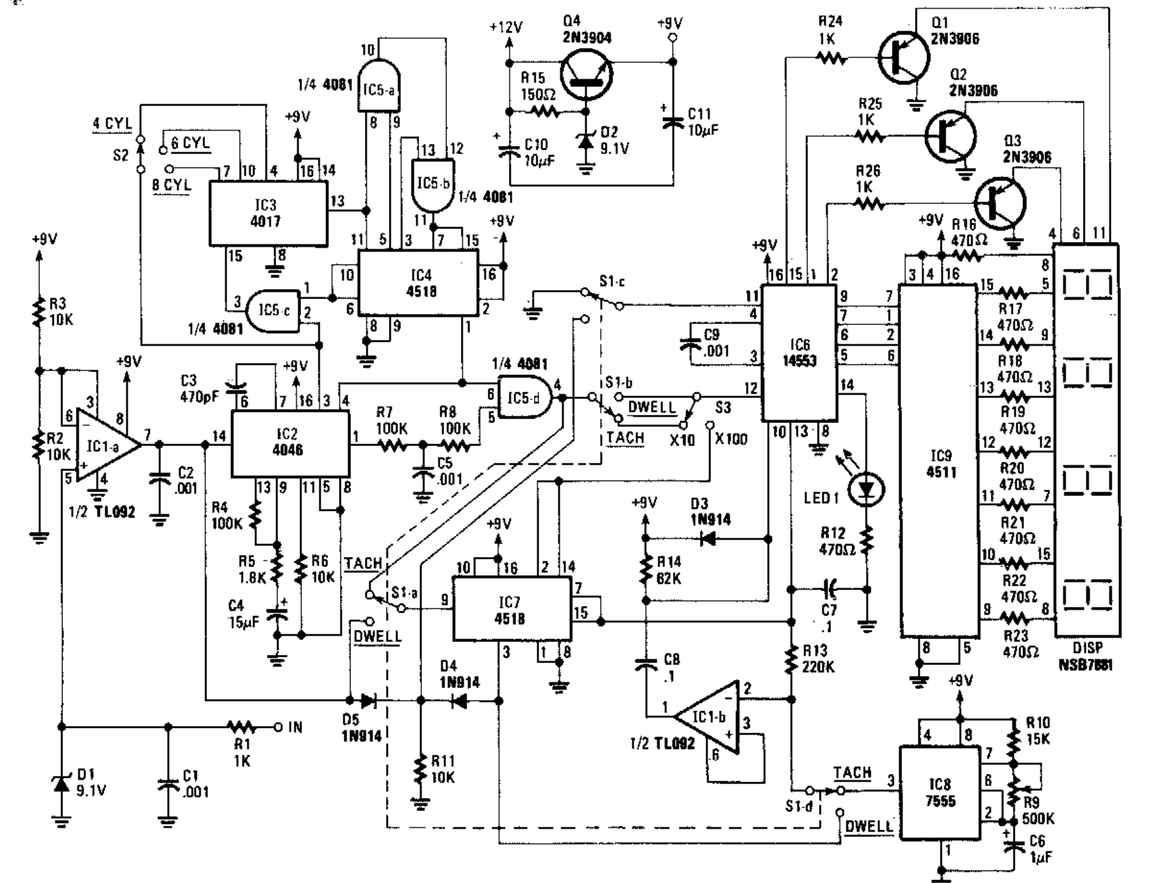


ZeTeX, formerly FERRANTI

Fig. 8-6

Operating the brake pedal of the car brings on the normal brake lights and then, after a delay, the extra lights are turned on. A bimetal strip in series with the lights would make them flash.

DIGITAL TACH/DWELL METER



Reprinted with permission from Radio-Electronics Magazine, July 1985. Copyright Gernsback Publications, Inc., 1985.

Fig. 8-7

The heart of the circuit is IC2, a 4046 micropower phase-locked loop (PLL). The incoming signals are fed to the PLL after being buffered by IC1a and its associated components. The frequency of the incoming signal is multiplied by either 90, 60, or 45, depending on the setting of the cylinder select switch, S2. That switch selects the proper output from counters IC3 and IC4, which are set to divide the output frequency of the PLL by those amounts, and then send the divided output back to the comparator to the PLL to keep it locked on to the input signal. The phase pulses output at pin 4 of IC2, then go through an AND gate IC5d—which only passes the signals if the PLL is locked on to an input signal, preventing stray readings—and then to the input of IC6. When in the tach mode, IC6 counts the number of pulses present at pin 12, during the timing interval generated by IC8 and the associated circuitry of IC1b. Because of the varied multiplication rate for the different cylinder selections—90, 60, and 45 for 4, 6, and 8 cylinders, respectively, the time interval is always constant at 1/3 of a second. The time interval is adjusted with R9, a 500-KΩ potentiometer; it is the only adjustment in the circuit.

In the high-tach (TACH 1 or ×100) range of 0 – 9990 rpm, the output of IC2 is routed by switches S1a and S3 through IC7, a divide-by-ten counter, which increases the count range tenfold. In the low-tach (TACH 2 or ×10) range of 0 – 999 rpm, the counter is bypassed.

AUTOMOBILE AIR CONDITIONER SMART CLUTCH

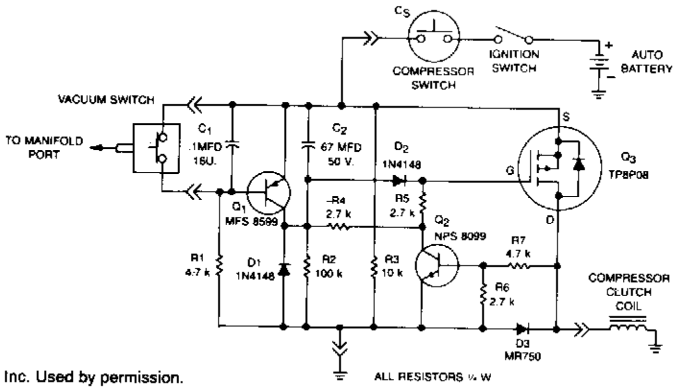


Fig. 8-8

This circuit disables the air conditioner compressor when additional engine power is required. It does so by monitoring the engine vacuum at the intake manifold. If the vacuum drops to 40% of its normal level, the compressor clutch is disabled, removing the air conditioner load from the engine. After the engine returns to normal vacuum level, there is a 6 second delay before the compressor clutch is enabled and the air conditioner is reactivated. This allows 6 seconds of extra power, about 500 ft at 60 MPH, which increases the safety margin when passing another vehicle. Loss of cooling is minimal because the air conditioner fan is not interrupted. When the engine is accelerated, manifold vacuum drops and vacuum switch VS opens to 40% of the normal manifold pressure. This causes Q1 to turn on, discharging C2 and turning off Q3 via diode D2. When Q3 turns off, so does Q2. When the engine reaches its normal operating vacuum, VS closes and Q1 turns off, allowing C2 to charge for 6 seconds until Q3 turns on again.

DOOR AJAR MONITOR

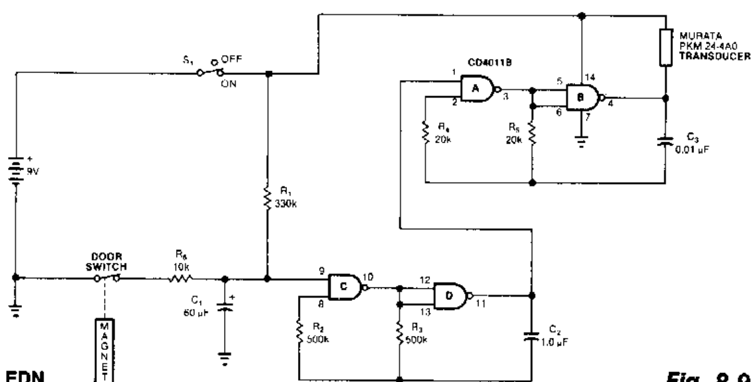
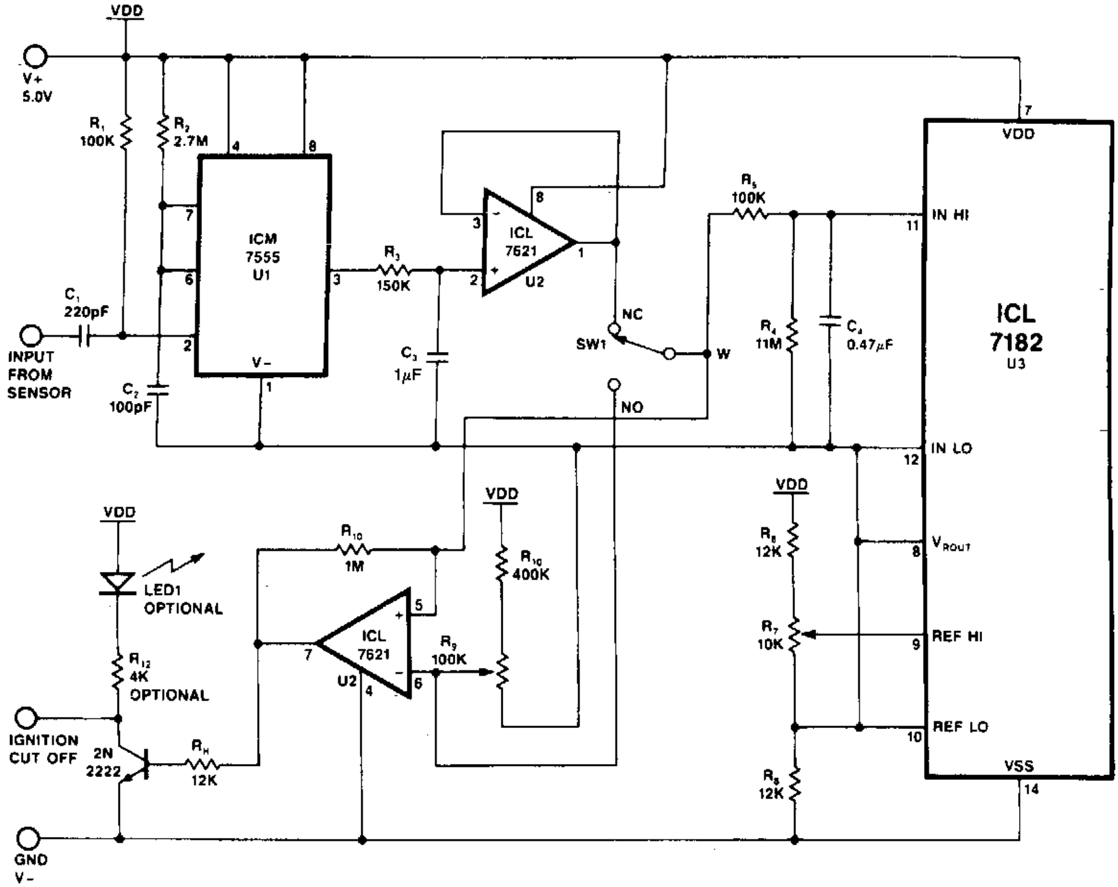


Fig. 8-9

The monitor senses an ajar door and, if the situation isn't corrected within 20 seconds, sounds a beeping alarm. The circuit is controlled by a magnetic reed switch and magnet on the door. With the door closed, the switch is closed and the alarm is disarmed. Opening the door opens switch, C1 starts charging up through R1. Approximately 20 seconds later, the voltage at pin 9 is high enough to turn on the oscillator formed from *C*, *D*, R2, R3, and C2. That pulses the piezoelectric transducer's 3-kHz oscillator. For lower standby drain on the battery, change R1 to 66 MΩ and C1 to 1 mF (film).

TACHOMETER WITH SET POINT



$$\frac{V_O}{V_{IN}} = \frac{1/RC}{S + 1/R_3C_3}$$

$$fc = \frac{1}{2\pi R_3 C_3}$$

SW1 Momentary
Switch SPST

V_{IN} = 264 mV @ 5000 RPM
4 Stroke V8

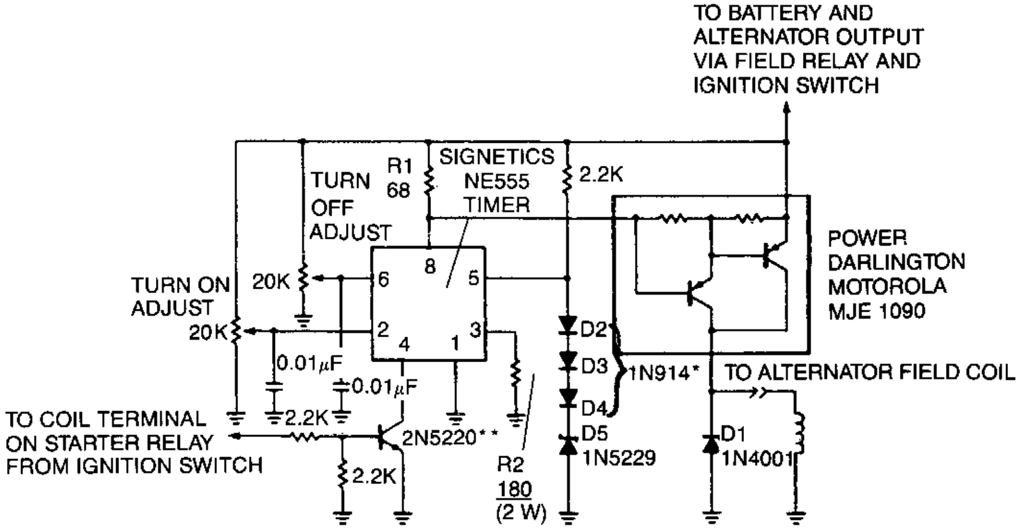
RPM	Hz	Period
600	10	100 ms
1000	16.7	60 ms
5000	83	12 ms
10,000	166.7	6 ms

No. of Cylinders	Events Per Cycle	Strokes Per Cycle
1	0.5	4
4	2	4
6	3	4
8	4	4

INTERSIL

Fig. 8-10

AUTOMOBILE VOLTAGE REGULATOR



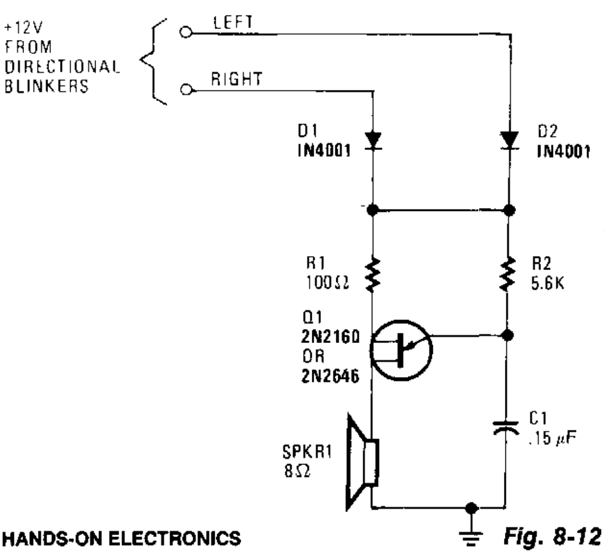
NOTES:
* Can be any general purpose Silicon diode or 1N4157 device.
**Can be any general purpose Silicon transistor.
All resistor values are in ohms.

SIGNETICS

Fig. 8-11

A monolithic 555-type timer is the heart of this simple automobile voltage regulator. When the timer is off so that its output at pin 3 is low, the power Darlington transistor pair is off. If battery voltage becomes too low, less than 14.4 V in this case, the timer turns on and the Darlington pair conducts.

DIRECTIONAL SIGNALS MONITOR

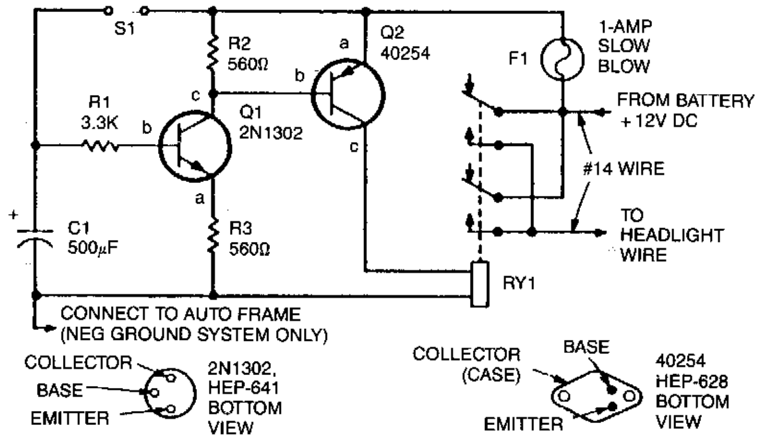


A unijunction transistor audio oscillator drives a small speaker. The oscillator's frequency is determined by resistor R2 and capacitor C2. The operating voltage is supplied from the car's turn-signal circuit(s) through D1 and D2. The diodes conduct current from the blinker circuit that is energized, and prevent stray current flow to the other blinker circuit.

HANDS-ON ELECTRONICS

Fig. 8-12

AUTOMATIC HEADLIGHT DELAY



C1—500 μF electrolytic capacitor, 15 VDC or better
Q1—NPN transistor, 2N1302, HEP-641
Q2—PNP, transistor, RCA 40254, HEP-628
R1—3,300-ohm, 1/2-watt resistor
R2, R3—560-ohm, 1/2-watt resistor
RY1—Relay, DPDT, 10-amp contacts, 12 VDC coil resistance at least 100-ohms, Potter and Brumfield type MR11D or equiv.

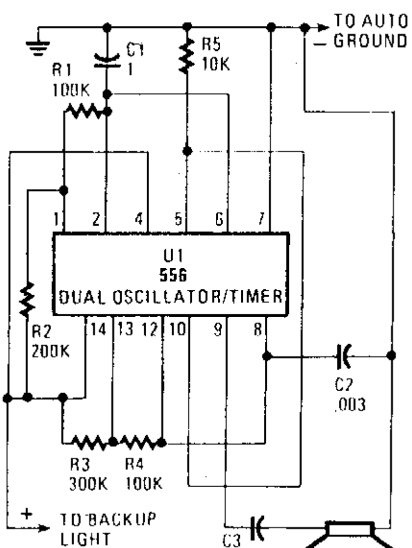
S1—SPST pushbutton switch, normally open (time-start switch)
Misc—3 1/4-in. × 2 1/8-in. × 1 5/8-in. case, 1-amp SB fuse with pigtail leads, #14 wire, hook-up wire, printed circuit material, hardware, solder, etc.

TAB BOOKS

Fig. 8-13

When the driver depresses pushbutton switch S1, timing capacitor C1 charges to 12 V and turns on transistor Q1, which drives power transistor Q2 into conduction. This, in turn, energizes the relay which has its contacts connected in parallel with the headlight switch. The relay will stay energized until C1 discharges to the Q1 turn-off level. The lights-on period is determined by the value of C1, R1, and the characteristics of transistor Q1. With values chosen on the schematic, about 60 light-on seconds are provided.

BACK-UP BEEPER

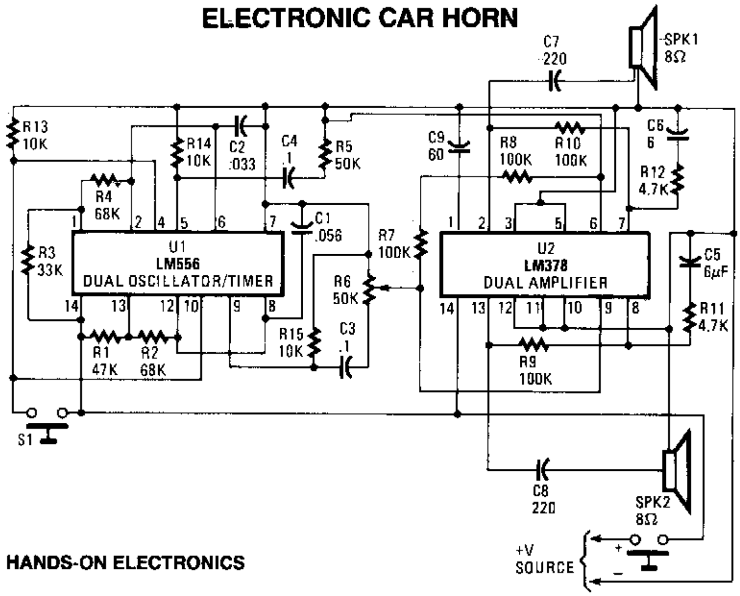


TAB BOOKS

Fig. 8-14

Put the car in reverse and the circuit provides a loud, audible beep at the rate of about one per second (1 Hz). Half of U1, a 556 dual oscillator/timer, is used as a slow-pulse oscillator with a rate of about 1 Hz. Components R2, R1, and C1 form the long time constant. You can calculate on time by t = .7 (R1 + R2) C1 or 1.15 seconds. The off time is shorter than the on time, at .77 second. Enabling pin 4 (reset) is held high to keep the oscillator free-running when voltage is applied to pin 14. The output at pin 5 is coupled to pin 10 of U1 enabling oscillator 2. Oscillator 2 of U1 produces an audio output of about 1 kHz, as determined by C2, R3, and R4. Pin 10 (reset) of oscillator 2 is connected to the pin 5 output of oscillator 1. So when pin 5 becomes positive, the oscillator beeps a short pulsed tone of 1 kHz.

ELECTRONIC CAR HORN

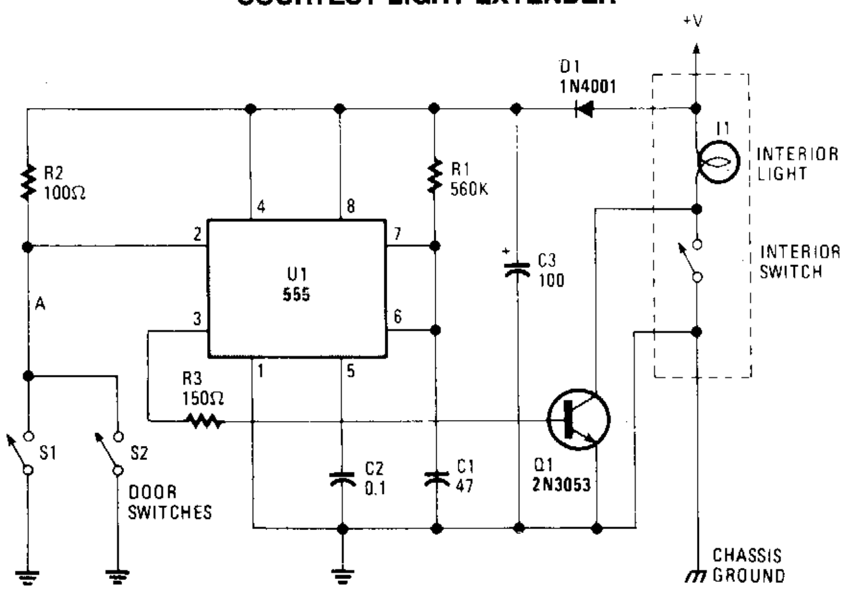


HANDS-ON ELECTRONICS

Fig. 8-15

An LM556 dual oscillator/timer, U1, configured as a two-tone oscillator drives U2, a dual 4-watt amplifier. One of the oscillators, pins 1 to 6, contained in U1 produces the upper frequency signal of about 200 Hz, while the second oscillator, pins 8 to 13, provides the lower frequency signal of about 140 Hz. Increase or decrease the frequencies by changing the values of C2 and C3. U1's outputs, pins 9 and 5, are connected to separate potentiometers to provide control over volume and balance. Each half of U2 produces 4 W of audio that is delivered to two 8-Ω loudspeakers via capacitors C7 and C8.

COURTESY LIGHT EXTENDER



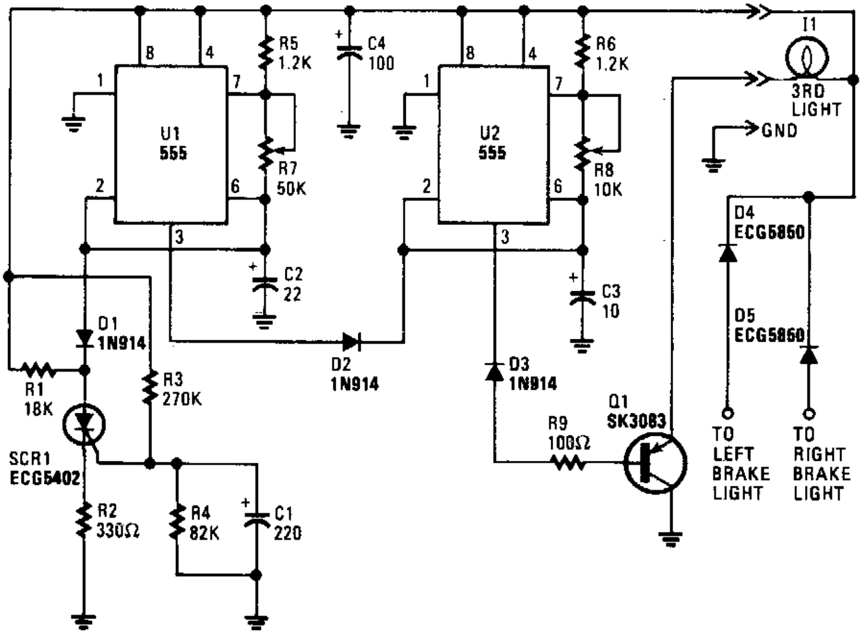
HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 8-16

COURTESY LIGHT EXTENDER (*Cont.*)

The circuit keeps the courtesy light on for 30 seconds after you close the door. The lead from the door switch is removed and connected to the 555 circuit. The 555 is arranged in a monostable mode, and is triggered by the door switches. The output drives Q1, which is connected across the interior light switch. The interior light is turned on for 30 seconds after the door is opened. If the door(s) are held open for longer than 30 seconds, it will not reset until after the doors are closed. In that case, the lights go out immediately.

FLASHING THIRD BRAKE LIGHT



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 8-17*

When power is first applied, three things happen: light-driving transistor Q1 is switched on due to a low output from U2 pin 3; timer U1 begins its timing cycle, with the output, pin 3, becoming high, inhibiting U2's trigger, pin 2, via D2; and charge current begins to move through R3 and R4 to C1.

When U1's output becomes low, the inhibiting bias on U2 pin 2 is removed, so U2 begins to oscillate, flashing the third light via Q1, at a rate determined by R8, R6, and C3. That oscillation continues until the gate-threshold voltage of SCR1 is reached, causing it to fire and pull U1's trigger, pin 2,low.

With its trigger low, U1's output is forced high, disabling U2's triggering. With triggering inhibited, U2's output switches to a low state, which makes Q1 conduct, turning on I1 until the brakes are released. Of course, removing power from the circuit resets SCR1, but the rc network consisting of R4 and C1 will not discharge immediately and will trigger SCR1 earlier. So, frequent brake use means fewer flashes.

HEADLIGHT ALARM

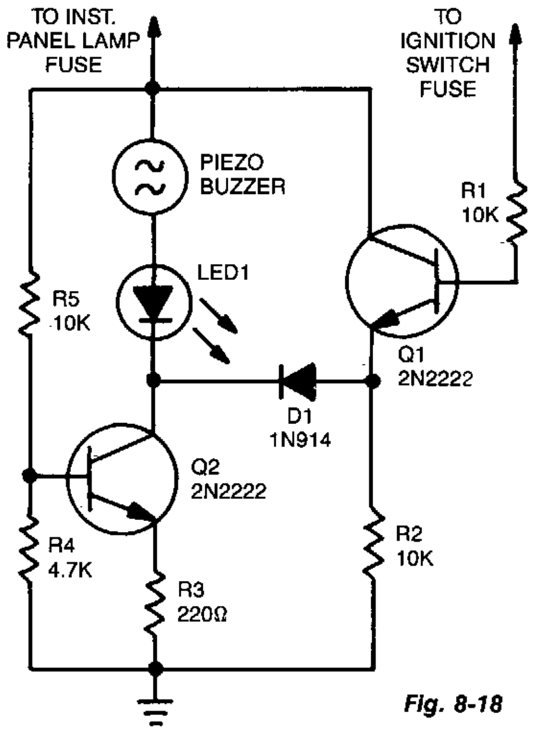


Fig. 8-18

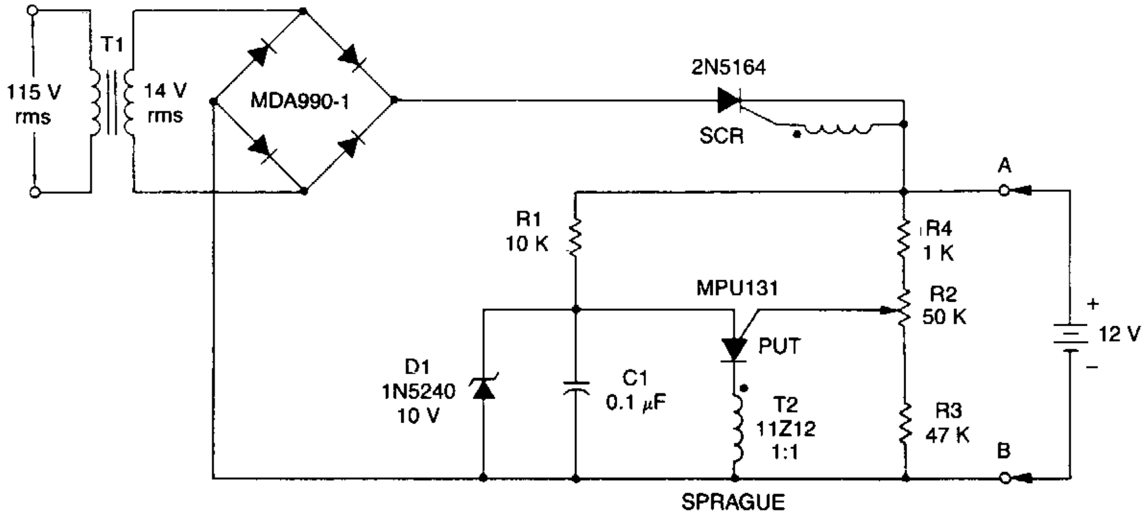
The base of Q1 is connected to the car's ignition circuit. One side of the piezoelectric buzzer is connected to the instrument-panel light fuse. When the headlights are off, no current reaches the buzzer, and therefore nothing happens. What happens when the headlights are on depends on the state of the ignition switch. When the ignition switch is on, transistors Q1 and Q2 are biased on, removing the buzzer and the LED from the circuit. When the ignition switch is turned off, but the headlight switch remains on; transistor Q1 is turned off, but transistor Q2 continues to be biased on. The result is that the voltage is sufficient to sound the buzzer loudly and light the LED. Turning off the headlight switch will end the commotion quickly.

9

Battery Chargers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PUT BATTERY CHARGER



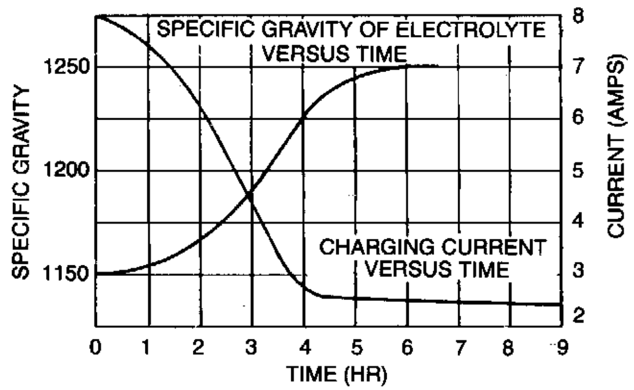
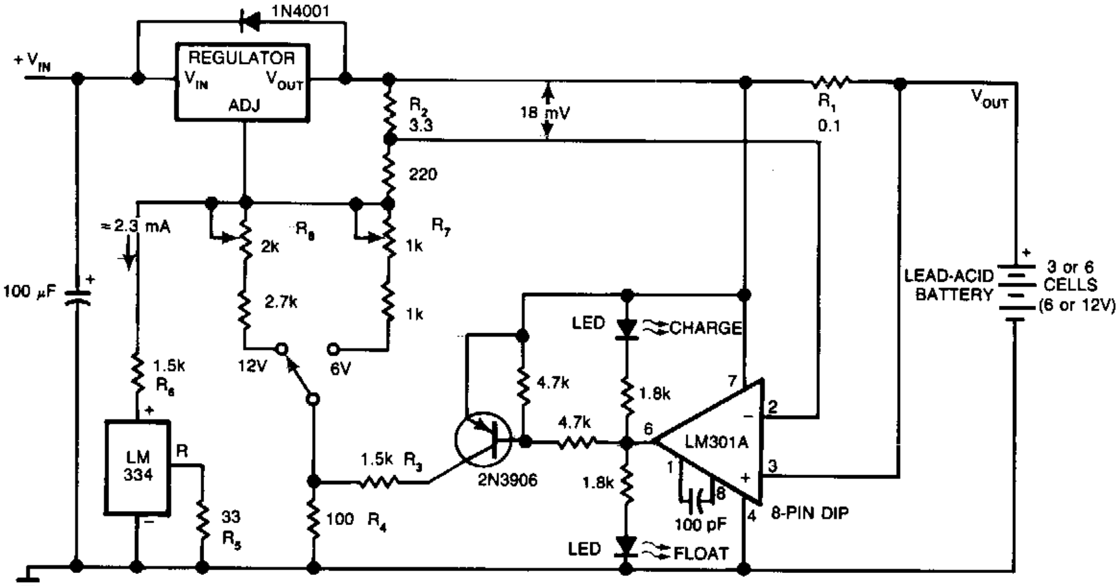


Fig. 9-1

A short-circuit-proof battery charger will provide an average charging current of about 8 A to a 12-V lead/acid storage battery. The charger circuit has an additional advantage; it will not function nor will it be damaged by improperly connecting the battery to the circuit. With 115 V at the input, the circuit commences to function when the battery is properly attached. The battery provides the current to charge the timing capacitor C1 used in the PUT relaxation oscillator. When C1 charges to the peak point voltage of the PUT, the PUT fires turning the SCR on, which in turn applies charging current to the battery. As the battery charges, the battery voltage increases slightly which increases the peak point voltage of the PUT. This means that C1 has to charge to a slightly higher voltage to fire the PUT. The voltage on C1 increases until the zener voltage of D1 is reached, which clamps the voltage on C1, and thus prevents the PUT oscillator from oscillating and charging ceases. The maximum battery voltage is set by potentiometer R2 which sets the peak point firing voltage of the PUT. In the circuit shown, the charging voltage can be set from 10 V to 14 V—the lower limit being set by D1 and the upper limit by T1.

LEAD/ACID BATTERY CHARGER



EDN

Fig. 9-2

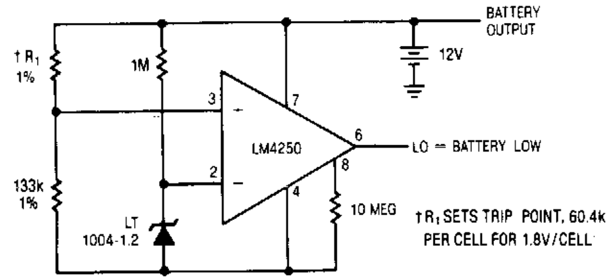
This circuit furnishes an initial voltage of 2.5 V per cell at 25°C to rapidly charge a battery. The charging current decreases as the battery charges, and when the current drops to 180 mA, the charging circuit reduces the output voltage to 2.35 V per cell, leaving the battery in a fully charged state. This lower voltage prevents the battery from overcharging, which would shorten its life.

The LM301A compares the voltage drop across R1 with an 18 mV reference set by R2. The comparator's output controls the voltage regulator, forcing it to produce the lower float voltage when the battery-charging current, passing through R1, drops below 180 mA. The 150 mV difference between the charge and float voltages is set by the ratio of R3 to R4. The LEDs show the state of the circuit.

Temperature compensation helps prevent overcharging, particularly when a battery undergoes wide temperature changes while being charged. The LM334 temperature sensor should be placed near or on the battery to decrease the charging voltage by 4 mV/°C for each cell. Because batteries need more temperature compensation at lower temperatures, change R5 to 30 Ω for a tc of −5 mV/°C per cell if application will see temperatures below −20°C.

The charger's input voltage must be filtered dc that is at least 3 V higher than the maximum required output voltage: approximately 2.5 V per cell. Choose a regulator for the maximum current needed: LM371 for 2 A, LM350 for 4 A, or LM338 for 8 A. At 25°C and with no output load, adjust R7 for a V_{OUT} of 7.05 V, and adjust R8 for a V_{OUT} of 14.1 V.

LEAD/ACID LOW-BATTERY DETECTOR



LINEAR TECHNOLOGY CORP.

Fig. 9-3

UNIVERSAL BATTERY CHARGER

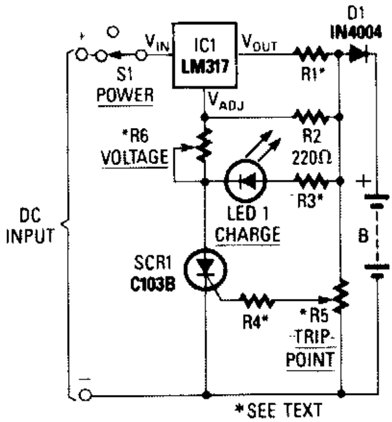
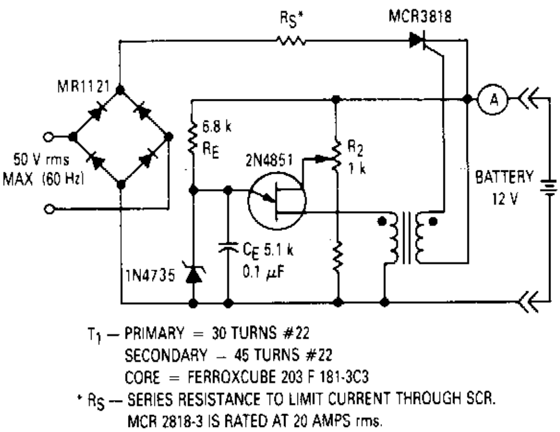


Fig. 9-4

Reprinted with permission from Radio-Electronics Magazine, July 1986. Copyright Gernsback Publications, Inc. 1986.

When power is applied to the circuit, SCR1 is off, so there is no bias-current path to ground; thus, LM317 acts as a current regulator. The LM317 is connected to the battery through steering diode D1, limiting resistor R1, and bias resistor R2. The steering diode prevents the battery from discharging through the LED and the SCR when power is removed from the circuit. As the battery charges, the voltage across trip-point potentiometer R5 rises, and at some point, turns on the SCR. Then, current from the regulator can flow to ground, so the regulator now functions in the voltage mode. When the SCR turns on, it also provides LED1 with a path to ground through R3. So, when LED1 is on, the circuit is in the voltage-regulating mode; when LED1 is off, the circuit is in the current-regulating mode.

UJT BATTERY CHARGER



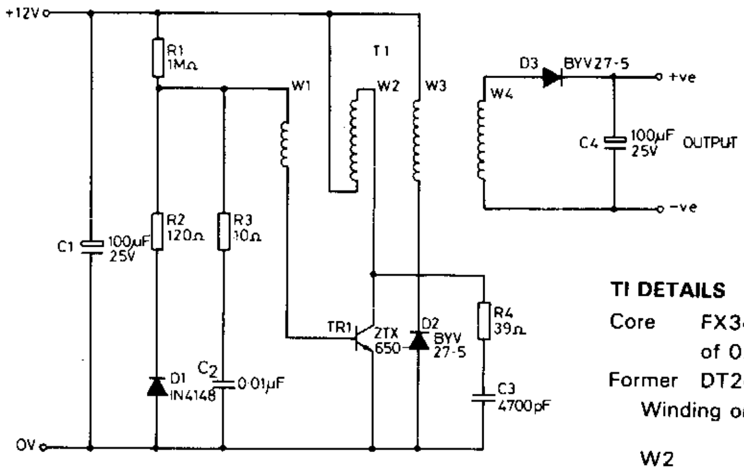
T1 — PRIMARY = 30 TURNS #22
SECONDARY — 45 TURNS #22
CORE = FERROXCUBE 203 F 181-3C3
* RS — SERIES RESISTANCE TO LIMIT CURRENT THROUGH SCR. MCR 2818-3 IS RATED AT 20 AMPS rms.

Copyright of Motorola, Inc. Used by permission.

Fig. 9-5

This circuit will not work unless the battery to be charged is connected with proper polarity. The battery voltage controls the charger and when the battery is fully charged, the charger will not supply current to the battery. The battery charging the current is obtained through the SCR when it is triggered into the conducting state by the UJT relaxation oscillator. The oscillator is only activated when the battery voltage is low. V_{B2B1} of the UJT is derived from the voltage of the battery to be charged, and since $V_P = V_D = V_{B2B1}$; the higher V_{B2B1}, the higher V_P. When V_P exceeds the breakdown voltage of the zener diode Z1, the UJT will cease to fire and the SCR will not conduct. This indicates that the battery has attained its desired charge as set by R2.

PORTABLE NICAD BATTERY CHARGER



T1 DETAILS

Core FX3437 With Gap/Spacer of 0.08mm

Former DT2492

Winding order W2, W4, W3 then W1

W2	40T	30awg.
W4	20T	30awg.
W3	13T	36awg.
W1	12T	36awg.

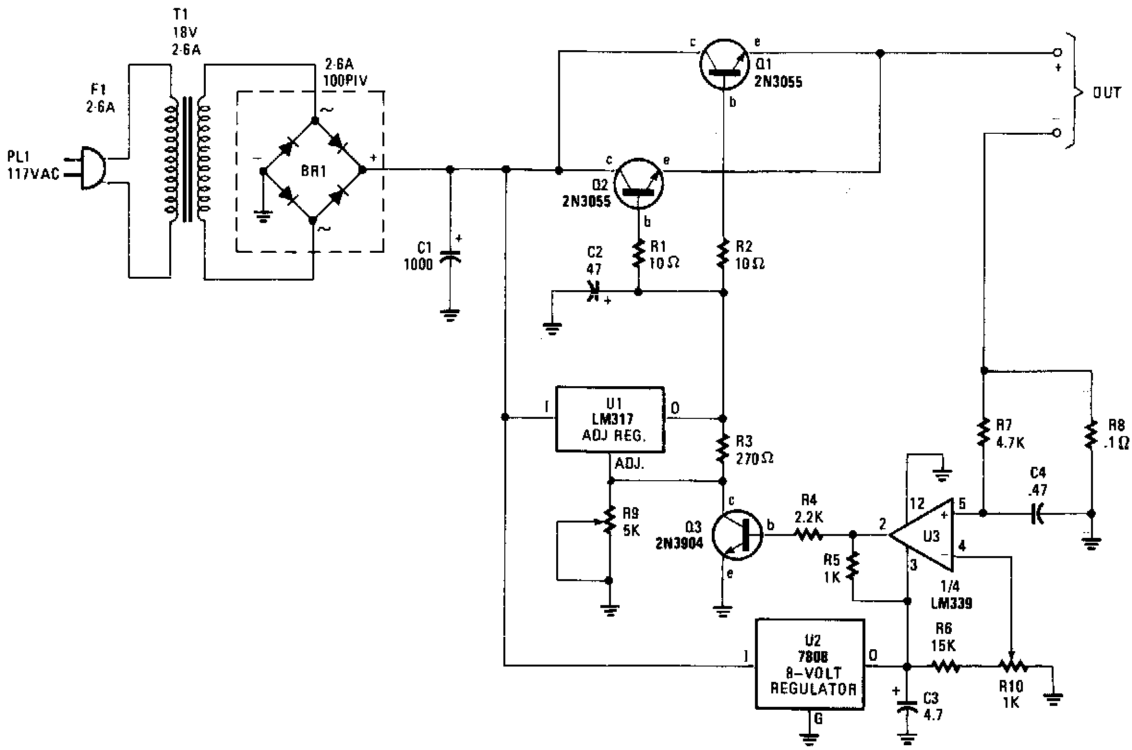
ZeTeX, formerly FERRANTI

Fig. 9-6

This circuit was designed to charge NiCad battery packs in the range of 4.8 to 15.6 V from a convenient remote power source, such as an automobile battery. When power is first applied to the circuit, a small bias current supplied by R1 via winding W1, starts to turn on the transistor TR1. This forces a voltage across W2 and the positive feedback given by the coupling of W1 and W2 causes the transistor to turn hard on, applying the full supply across W2. The base drive voltage induced across W1 makes the junction between R1 and R2 become negative with respect to the 0-V supply, forward-biasing diode D1 to provide the necessary base current to hold TR1 on.

With the transistor on, a magnetizing current builds up in W2, which eventually saturates the ferrite core of transformer T1. This results in a sudden increase on the collector current flowing through TR1, causing its collector-emitter voltage to rise, and thus reducing the voltage across W2. The current flowing in W2 forces the collector voltage of the TR1 to swing positive until restricted by transformer output loading. Rc network R4 and C3 limits the turn off transient TR1. R3 and C2 maintain the loop gain of the circuit when diode D1 is not conducting.

UNIVERSAL BATTERY CHARGER



POPULAR ELECTRONICS

Fig. 9-7

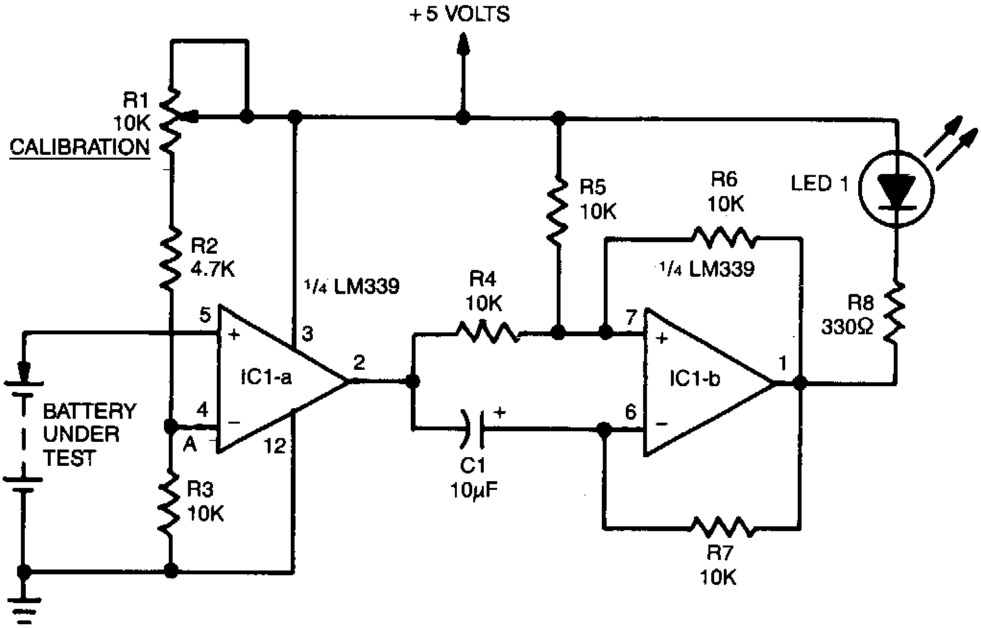
The charger's output voltage is adjustable and regulated, and has an adjustable constant-current charging circuit that makes it easy to use with most NiCad batteries. The charger can charge a single cell or a number of series-connected cells up to a maximum of 18 V.

Power transistors Q1 and Q2 are connected as series regulators to control the battery charger's output voltage and charge-current rate. An LM317 adjustable voltage regulator supplies the drive signal to the bases of power transistors Q1 and Q2. Potentiometer R9 sets the output-voltage level. A current-sampling resistor, R8 (a 0.1-Ω, 5-W unit), is connected between the negative output lead and circuit ground. For each amp of charging that flows through R8, a 100 mV output is developed across it. The voltage developed across R8 is fed to one input of comparator U3. The other input of the comparator is connected to variable resistor R10.

As the charging voltage across the battery begins to drop, the current through R8 decreases. Then the voltage feeding pin 5 of U3 decreases, and the comparator output follows, turning Q3 back off, which completes the signal's circular path to regulate the battery's charging current.

The charging current can be set by adjusting R10 for the desired current. The circuit's output voltage is set by R9.

LOW-BATTERY WARNING



GERNSBACK PUBLICATIONS INC.

Fig. 9-8

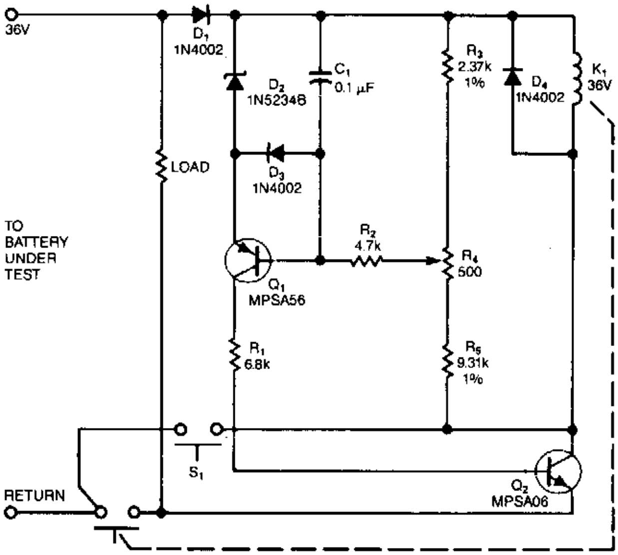
A voltage divider consisting of R1, R2, and R3 is used to set the input reference voltage below which the batteries are to be replaced. That reference voltage, at point A, is varied by R1. With the voltage divider shown in Fig. 9-7, a range of 2 to 3.5 V is possible. When the battery voltage drops below that at point A, the output of IC1a, 1/4 of a LM339 quad comparator, switches from high to low. That triggers IC1b, which is configured as an astable multivibrator. Feedback resistors R6 and R7, coupled with capacitor C1, determine the time constant of the multivibrator. The output from IC1b is connected to LED1 through dropping resistor R8. With the circuit values as shown, the LED will flash at a rate of 3 Hz. Although this circuit was designed specifically to monitor RAM back-up batteries, it can of course be modified for use in just about any application where the condition of a battery must be found.

10

Battery Monitors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

QUICKLY DEACTIVATING BATTERY SENSOR



EDN *Fig. 10-1*

The sensing circuit rapidly disconnects the battery voltage and load whenever the voltage drops below a preset threshold. One-way operation prevents the circuit from reconnecting the load if the voltage should then rise above the threshold. C1 ensures that the circuit doesn't activate while making connections to the battery; if you accidentally reverse these connections, D1 will block the turn on the relay.

After you connect the battery, nothing happens until you depress pushbutton switch S1, which allows relay K1 to energize. When you release S1, the relay remains on only if the battery voltage is above the minimum level. You preset this threshold—to 31.5 V when testing 36-V batteries, for example—using R4. Q1 begins to turn off as the battery voltage drops. Once the threshold level is reached, Q2 also begins to turn off, and its rising collector voltage provides positive feedback to the base of Q1, accelerating the turn off. When Q2 turns off, the relay drops out, disconnecting the battery from its load.

AUTOMATIC SHUTOFF FOR BATTERY-POWERED PROJECTS

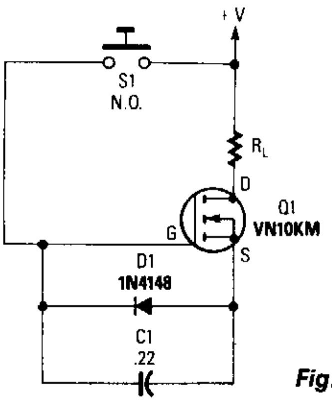
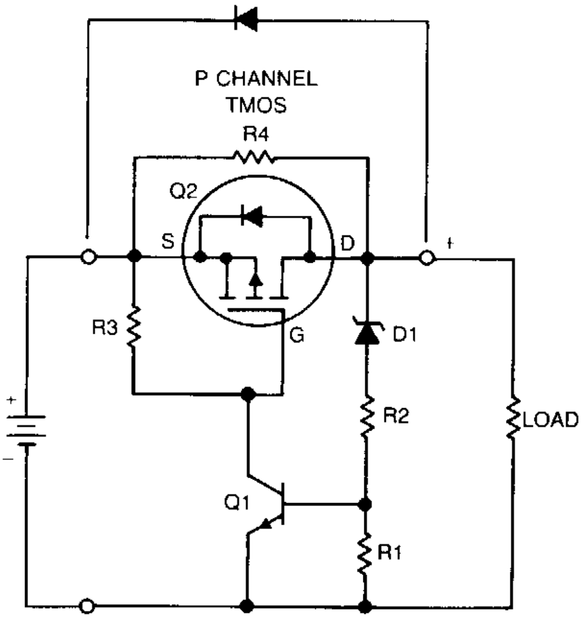


Fig. 10-2

POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

When S1 is depressed, C1 begins to charge to the supply voltage. That places a forward bias on the gate of Q1 turning it on and supplying current to load resistor RL. When the charge on C1 leaks off, the transistor shuts off, cutting off current to the load. That load could be anything from a transistor radio to a child's toy. Transistor Q1, available from Radio Shack as part No. 276–2070, is rated at 0.5 A at 60 Vdc. With a supply voltage of 9 Vdc and with C1 rated at 0.22 µF, a delay of about one minute is produced; with C1 rated at 10 µF, the delay is about an hour.

NICAD-BATTERY PROTECTION CIRCUIT



Copyright of Motorola, Inc. Used by permission.

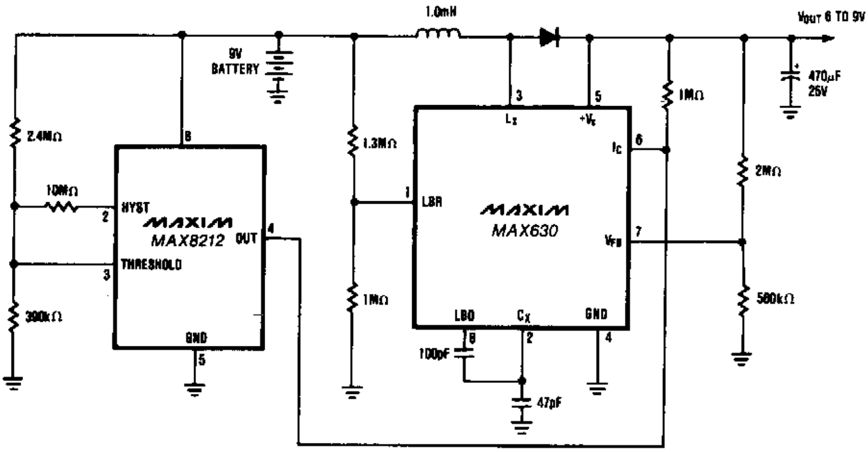
Fig. 10-3

If a NiCad battery is discharged to a point at which the lowest capacity cell becomes fully discharged and reverses polarity, that cell will usually short internally and become unusable. To prevent this type of damage, this circuit detects a one-cell drop of 1.25 V and turns the load off before cell reversal can occur.

Low-current zener or other voltage sensor D1 and resistors R1 and R2 establish a reference level for transistor Q1. These resistors bias the zener to a few microamperes above its "knee." Therefore, if battery voltage falls more than 1.25 V, Q1 turns off, turning off Q2, and disconnecting the load. After the load is disconnected, if the battery returns to nominal voltage, the high value of resistor shunting Q2 provides enough output voltage to reset the voltage sensor and turn Q2 back on. If desirable, shunt diode D2 or the parasitic diode of the TMOS device, if suitable, allows the battery to be charged from the load terminals.

The protection circuit presents a shunt current of only 10 mA at nominal battery voltage, which is low relative to the internal leakage of the batteries.

9-V BATTERY LIFE EXTENDER



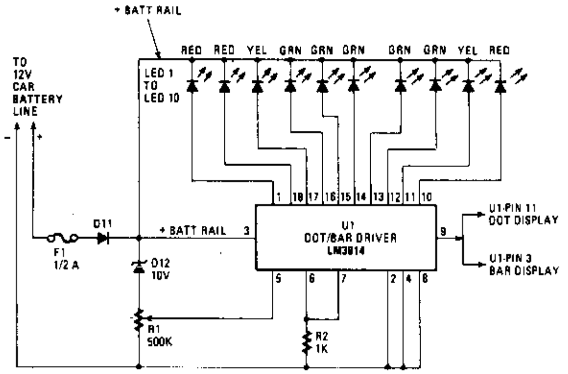
MAXIM

Fig. 10-4

9-V BATTERY LIFE EXTENDER (*CONT.*)

Circuit provides a minimum of 7 V until the 9-V battery voltage falls to less than 2 V. When the battery voltage is above 7 V, the MAX630's IC pin is low, putting it into the shutdown mode which draws only 10 nA. When the battery voltage falls to 7 V, the MAX8212 voltage detector's output increases. The MAX630 then maintains the output voltage at 7 V. The low battery detector (LBD) is used to decrease the oscillator frequency when the battery voltage falls to 3 V, thereby increasing the output current capability of the circuit. Note that this circuit, with or without the MAX8212, can be used to provide 5 V from 4 alkaline cells. The initial voltage is approximately 6 V, and the output is maintained at 5 V even when the battery voltage falls to less than 2 V.

AUTO BATTERY ALTERNATOR MONITOR

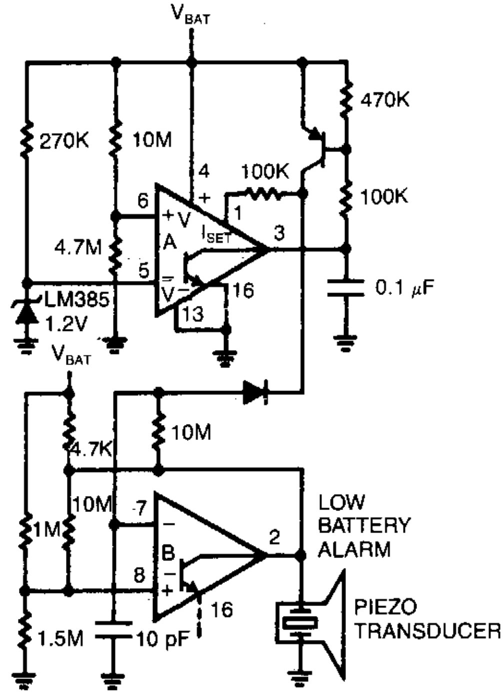


HANDS-ON ELECTRONICS *Fig. 10-5*

Most of the circuitry is contained in the LM3914 dot/bar-graph driver IC chip. In addition to the comparator circuitry within the package, it also contains a stable reference supply and the drivers for the LEDs. Resistor R2 acts as the current limiter for all the LEDs. Resistor R2 may be varied for LED brightness.

The unit will illuminate one LED for each voltage condition encountered in the charging system. This system is called a dot-graph display; it is achieved by wiring the mode control at pin 9 to pin 11 on U1. It is possible to wire the monitor so that each lamp will be illuminated up to the maximum voltage on the line at that moment. The latter is referred to as a bar-graph display. By connecting pin 9 to pin 3 on U1, the bar-graph mode will be enabled.

LOW-BATTERY DETECTOR

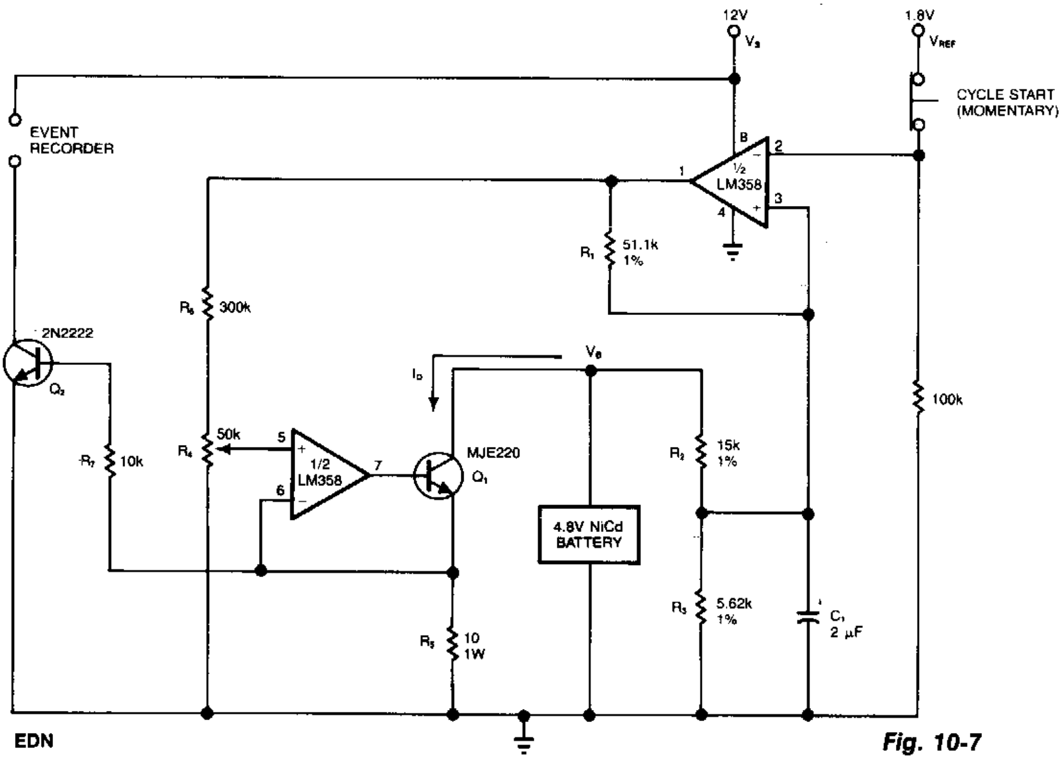


NATIONAL SEMICONDUCTOR CORP. *Fig. 10-6*

Comparator A detects when the supply voltage drops to 4 V and enables comparator B to drive a piezoelectric alarm.

I_S: 6 V at 45 μA
I_S: 3.8 V at 1 μA
f: 3 kHz

NICAD-BATTERY ANALYZER



EDN

Fig. 10-7

Because NiCad batteries maintain a constant output voltage, it is difficult to determine how much of the battery's charge remains. The circuit provides a way of determining the capacity of a battery by draining it at a preset current to its depleted voltage of 1 V/cell. Measure the discharge time of the cells and perform a simple calculation to obtain the battery's capacity.

Set the drain current (I_D) to $0.5C$ (C = battery capacity in mA/hr) by selecting an appropriate value for R4. Choose R5 such that: $I_D \times \text{R5} = 1$ V. V_{REF} is set so the comparator turns off the drain current and timer when the battery reaches its depleted voltage, V_B (usually 1 V/cell). You calculate V_{REF} as follows:

$$V_{REF} = \frac{R_3\,[R_2(V_S - 1.3) + R_1 V_B]}{R_1\,R_2 + R_2\,R_3 + R_1\,R_3}$$

With the battery in place, activate the circuit by grounding V_{REF} with the momentary switch. The battery drains at I_D until it reaches V_B, turning off the drain circuit and the timer. Hysteresis keeps the circuit from restarting. Determine the battery's capacity using the following equation:

$$C_{(\text{mAhr})} = \text{Time of Cycle} \times I_D$$

The circuit shown tests 4.8 V, 180 mA/hr batteries. I_D is 100 mA and V_B is 4 V.

LOW-BATTERY PROTECTOR

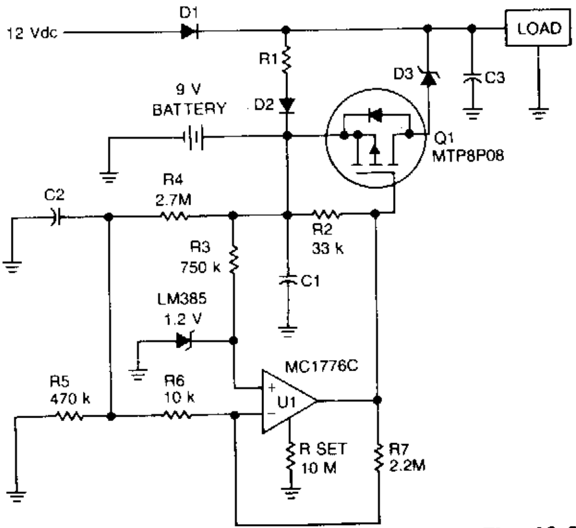
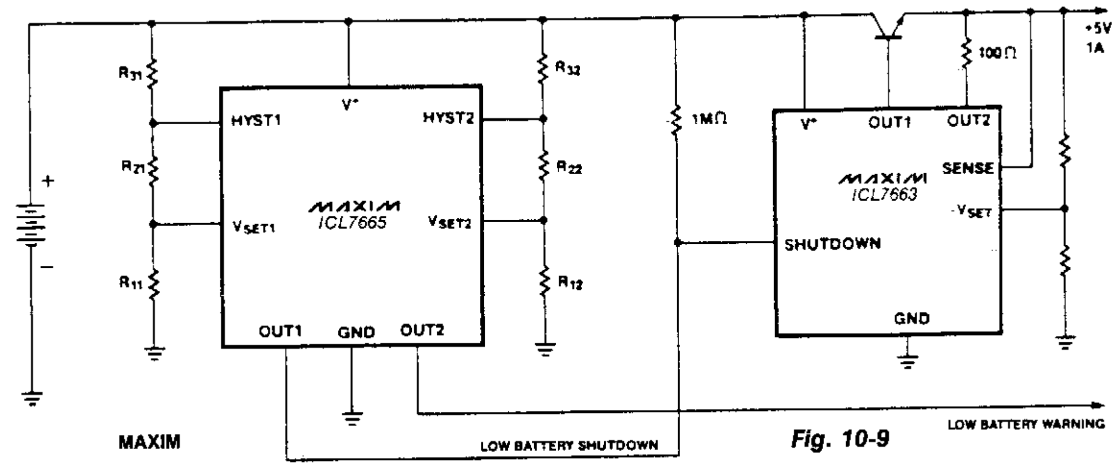


Fig. 10-8

To prevent battery damage due to over-discharge, a low-voltage detector and switch should be included in the design of the battery backup circuit. The detector circuit should consume extremely low current. The switch should exhibit a low-voltage drop and be easy to control.

R1 and D2 provide a trickle charge for the battery. Chosen for its low forward voltage drop, Schottky diode D3 prevents forward polarization of the diode incorporated in Q1. When the battery voltage is above approximately 8 V, the output of U1 is low and Q1 is turned on. If the battery voltage falls below 8 V, the output of U1 increases and turns off Q1.

LOW-BATTERY WARNING/DISCONNECTOR

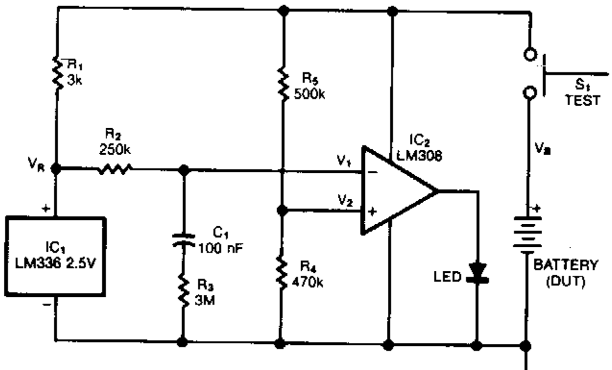


MAXIM

Fig. 10-9

NiCad batteries are excellent rechargeable power sources for portable equipment, but care must be taken to ensure that the batteries are not damaged by overdischarge. Specifically, a NiCad battery should not be discharged to the point where the polarity of the cell is reversed and is reverse-charged by the higher-capacity cells. This reverse charging will dramatically reduce the life of a NiCad battery. This circuit both prevents reverse charging and also gives a low-battery warning. A typical low-battery warning voltage is 1 V per cell. Since a NiCad, 9-V battery is ordinarily made up of six cells with a nominal voltage of 7.2 V, a low-battery warning of 6 V is appropriate, with a small hysteresis of 100 mV. To prevent overdischarge of a battery, the load should be disconnected when the battery voltage is 1 V × ($N-1$), where N = number of cells. In this case, the low-battery load disconnect should occur at 5 V. Since the battery voltage will rise when the load is disconnected, 800 mV of hysteresis is used to prevent repeated on-off cycling.

BATTERY CAPACITY TESTER



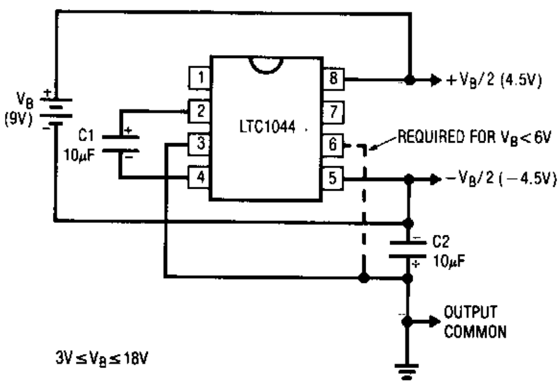
EDN

Fig. 10-10

The test circuit gives an indication of the capacity remaining in a battery. By noting the time in seconds that the LED remains on after you depress the test switch S1. The circuit has proven reliable in testing NiCad-, carbon-, and alkaline-type batteries. Closing S1 activates the circuit by applying voltage from the battery under test. Voltage V_1 jumps to a value $V_0 = V_R R_3/(R_2 + R_3)$ when the switch closes and then increases with a time constant $T = C_1 (R_2 + R_3)$. The divider R4/R5 fixes V_2. The reference circuit IC1 sets V_R to approximately 2.5 V. The op amp's output remains high (LED on) until V_1 rises to the level of V_2, when the LED turns off. Calculate the on-time t_{ON} as follows:

$$t_{ON} = T \ln \frac{V_R - V_0}{V_R - V_2}$$

BATTERY SPLITTER

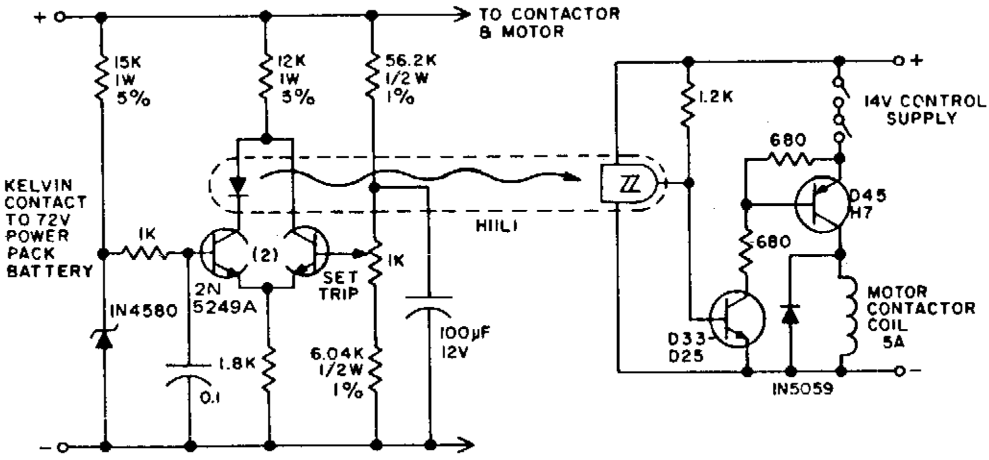


LINEAR TECHNOLOGY

Fig. 10-11

A common need in many systems is to obtain positive and negative supplies from a single battery. Where current requirements are small, the circuit shown is a simple solution. It provides symmetrical ± output voltages, both equal to one half the input voltage. The output voltages are referenced to pin 3, output common. If the input voltage between pin 8 and pin 5 exceeds 6 V, pin 6 should also be connected to pin 3, as shown by the dashed line. Higher current requirements are served by an LT1010 buffer. The splitter circuit can source or sink up to ±150 mA with only 5 mA quiescent current. The output capacitor, C2, can be made as large as necessary to absorb current transients. An input capacitor is also used on the buffer to avoid high frequency instability that can be caused by high source impedance.

ELECTRIC VEHICLE BATTERY SAVER



GE

Fig. 10-12

The battery life and operating cost of an electric vehicle is severely affected by overdischarge of the battery. This circuit provides both warning and shutdown. An electronic switch is placed in series with the propulsion motor contactor coil. Three modes of operation are possible:

- When the propulsion power pack voltage is above the 63-V trip point, the electronic switch has no effect on operation
- When the propulsion power pack no load voltage is below 63 V, power will not be supplied to the propulsion motor since the electronic switch will prevent contactor operation
- When the propulsion power pack loaded voltage drops below 63 V, the contactor will close and open because of the electronic switch. The ***bucking operation*** indicates to the operator need to charge the batteries

11

Bridge Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

AUTO-ZEROING SCALE

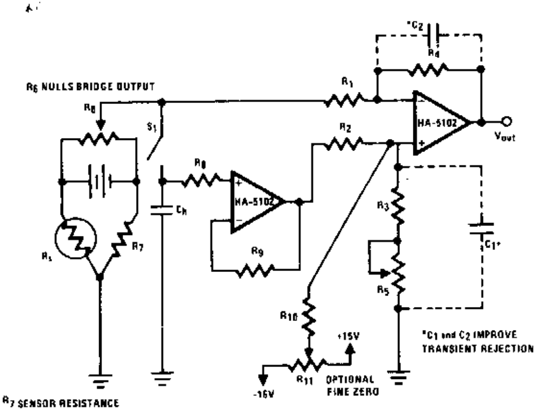
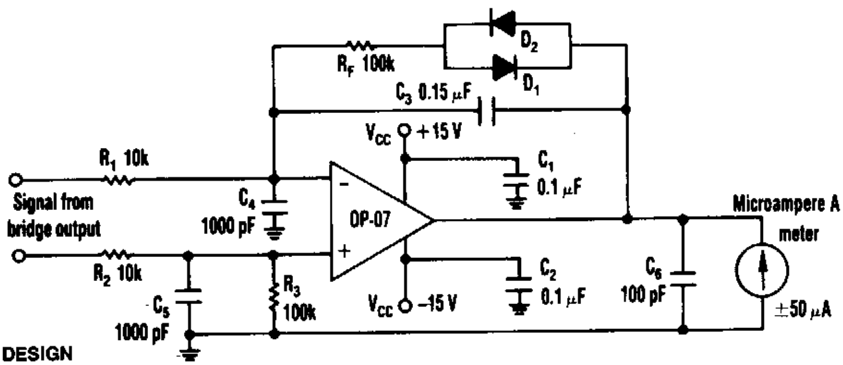


Fig. 11-1 **HARRIS**

Electronic scales have come into wide use and the HA-510X, as a very low noise device, can improve such designs. This circuit uses a stain-gauge sensing element as part of a resistive Wien-bridge. An auto-zero circuit is also incorporated into this design by including a sample-and-hold network.

The bridge signal drives the inverting input of a differentially configured HA-5102. The noninverting input is driven by the other half of the HA-5102 used as a buffer for the holding capacitor, *CH*. This second amplifier and its capacitor *CH* form the sampling circuit used for automatic output zeroing. The 20-KΩ resistor between the holding capacitor *CH* and the input terminal, reduces the drain from the bias currents. A second resistor *RG* is used in the feedback loop to balance the effect of R8. If R7 is approximately equal to the resistance of the strain gauge, the input signal from the bridge can be roughly nulled with R6. With very close matching of the ratio R4/R1 to R3/R2, the output offset can be nulled by closing S1. This will charge *CH* and provide a 0-V difference to the inputs of the second amplifier, which results in a 0-V output. In this manner, the output of the strain gauge can be indirectly zeroed. R10 and potentiometer R11 provide an additional mechanism for fine tuning V_{OUT}, but they can also increase offset voltage away from the zero point. C1 and C2 reduce the circuit's susceptibility to noise and transients.

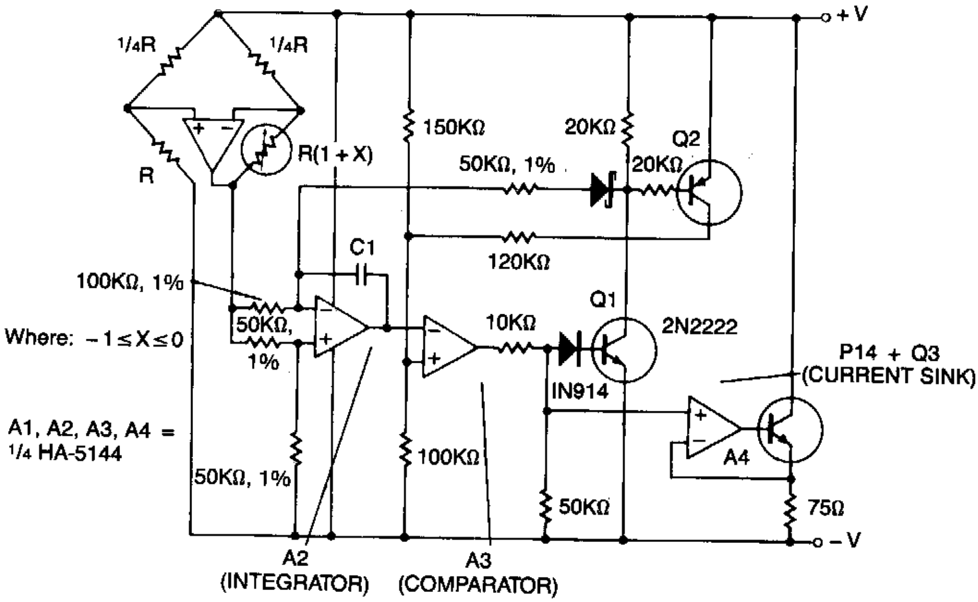
ACCURATE NULL/VARIABLE GAIN CIRCUIT



ELECTRONIC DESIGN *Fig. 11-2*

The circuit can use any general-purpose, low-offset, low-drift op amp, such as the OP-07. The differential signal from the bridge feeds an amplifier that drives an ordinary, rugged ±50-μA meter. Near the null point, however, the drastically reduced signal level from the bridge requires very high gain to achieve a high null resolution. To provide the variable-gain feature, the op amp's feedback path needs a dynamic resistance that increases as the input signal drops. Two common signal diodes, D1 and D2, in an antiparallel configuration in the feedback path supply function for all positive and negative inputs. To stabilize the op amp circuit at high gain, capacitors C3, C5, and C6 reduce response to high frequencies; capacitors C1 and C2 bypass the amplifier's power supplies.

REMOTE SENSOR LOOP TRANSMITTER



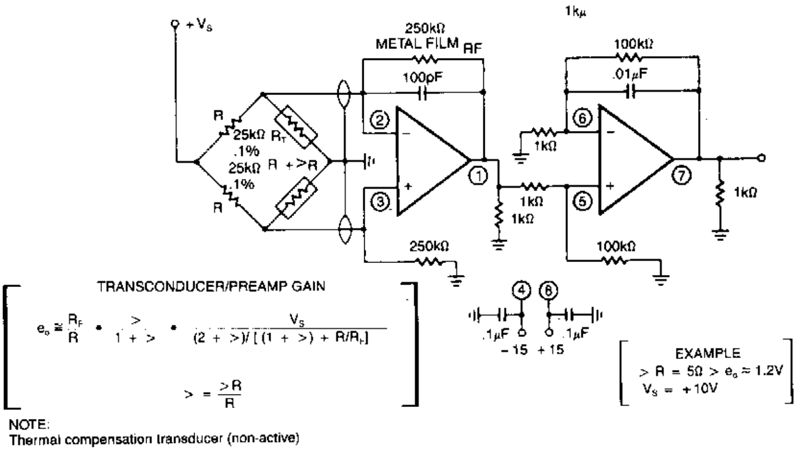
HARRIS

Fig. 11-3

This circuit shows amplifier A1 as a sensor amplifier in a bridge configuration. Amplifiers A2 and A3 are configured as a voltage to frequency converter and A4 is used as the transmitter. This entire sensor/transmitter can be powered directly from a 4 to 20 mA current loop.

The bridge configuration produces a linear output with respect to the changes in resistance of the sensor. The voltage at the output of A1 causes the integrator output A2 to ramp down until it crosses the comparator threshold voltage of A3. A3 turns on Q1 and Q2. A1 causes the output of A2 to ramp up at a rate nearly equal to its negative slope, while Q2 provides hysteresis for the comparator. In addition, Q1 and Q2 help eliminate changes in power supply loop voltage. Amplifier A4 and Q3 are configured as a constant current sink which turns on when the comparator current increases. The resulting increase in loop current transmits the frequency of the V/F converter back to the control circuitry.

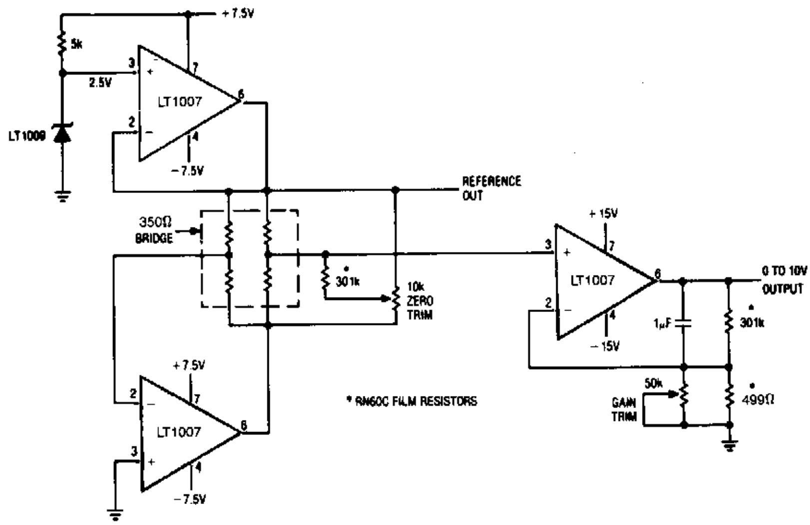
BRIDGE TRANSDUCER AMPLIFIER



SIGNETICS *Fig. 11-4*

In applications involving strain gauges, accelerometers, and thermal sensors, a bridge transducer is often used. Frequently, the sensor elements are high resistance units requiring equally high bridge resistance for good sensitivity. This type of circuit then demands an amplifier with high input impedance, low bias current and low drift. The circuit shown represents a possible solution to these general requirements.

STRAIN GAUGE SIGNAL CONDITIONER WITH BRIDGE EXCITATION



LINEAR TECHNOLOGY *Fig. 11-5*

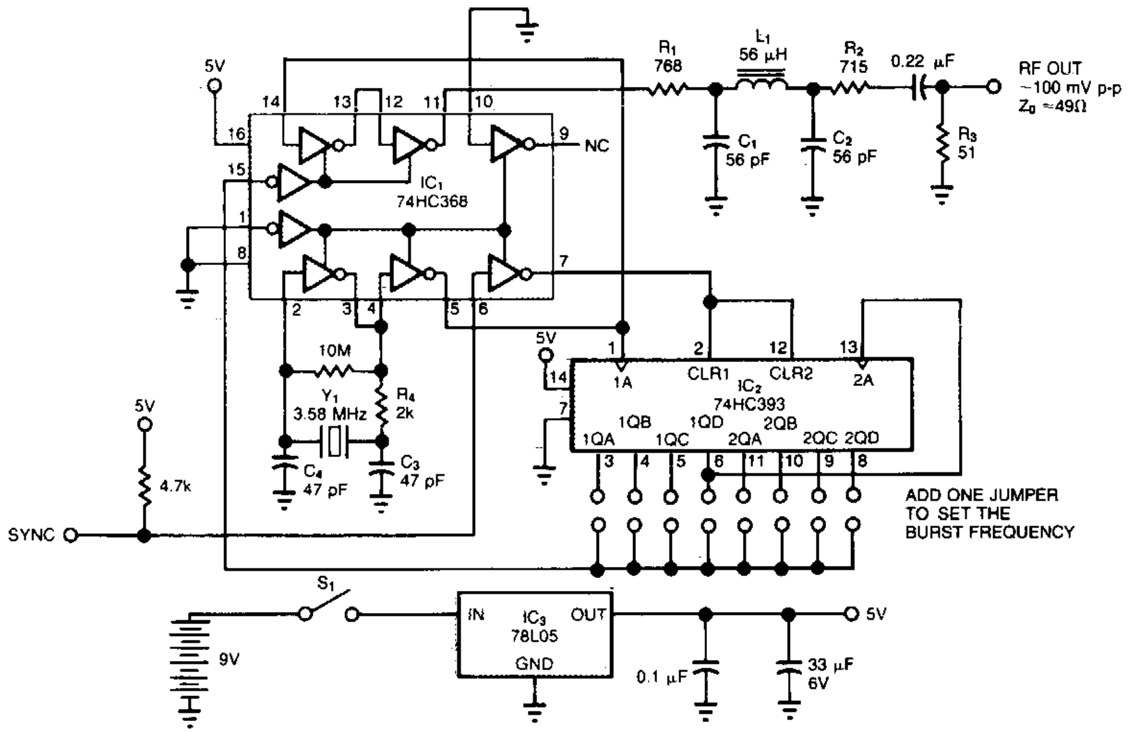
The LT1007 is capable of providing excitation current directly to bias the 350-Ω bridge at 5 V. With only 5 V across the bridge, as opposed to the usual 10 V, total power dissipation and bridge warm-up drift is reduced. The bridge output signal is halved, but the LT1007 can amplify the reduced signal accurately.

12

Burst Generators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PORTABLE RF BURST GENERATOR

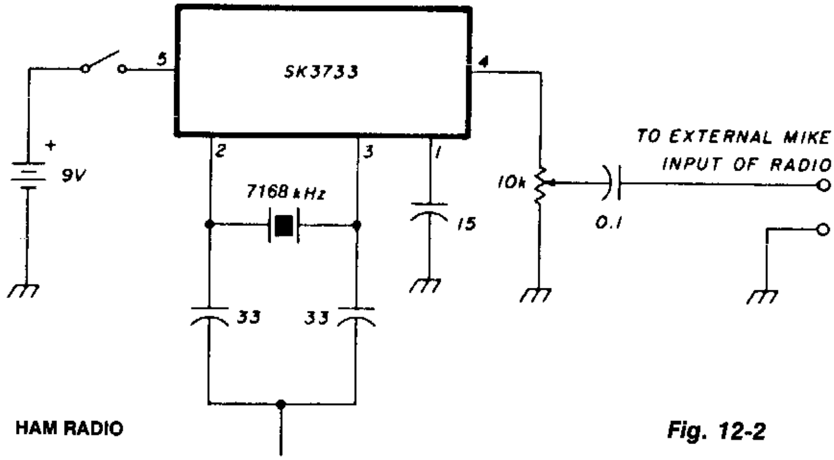


EDN

Fig. 12-1

The circuit generates low-level rf bursts having frequencies as high as 10 MHz, thus permitting field testing of high frequency receivers. A jumper-selectable binary fraction (1/2 to 1/256 of the Y1 crystal frequency gates the output rf signal. Output amplitude (open circuit) is approximately 100 mV; output impedance is approximately 49 Ω. The rf source is a clock oscillator based on a 3.58-MHz, color-burst crystal and two inverting buffers. The oscillator drives two cascaded 4-bit binary counters, IC2, and the sync signal resets the counters with a logic-high pulse—logic low at the counters. Select the desired output frequency by adding a jumper to one of the counter's eight output lines, which provides an enable signal for the two 3-state output buffers. The square-wave output at IC1, pin 11, is attenuated by R1, R2, and R3 to fix the output resistance at approximately 49 Ω. Resistor R3 is the only critical component; for clean gating, isolate it from the rest of the circuit.

TONE BURST GENERATOR FOR EUROPEAN REPEATERS



HAM RADIO

Fig. 12-2

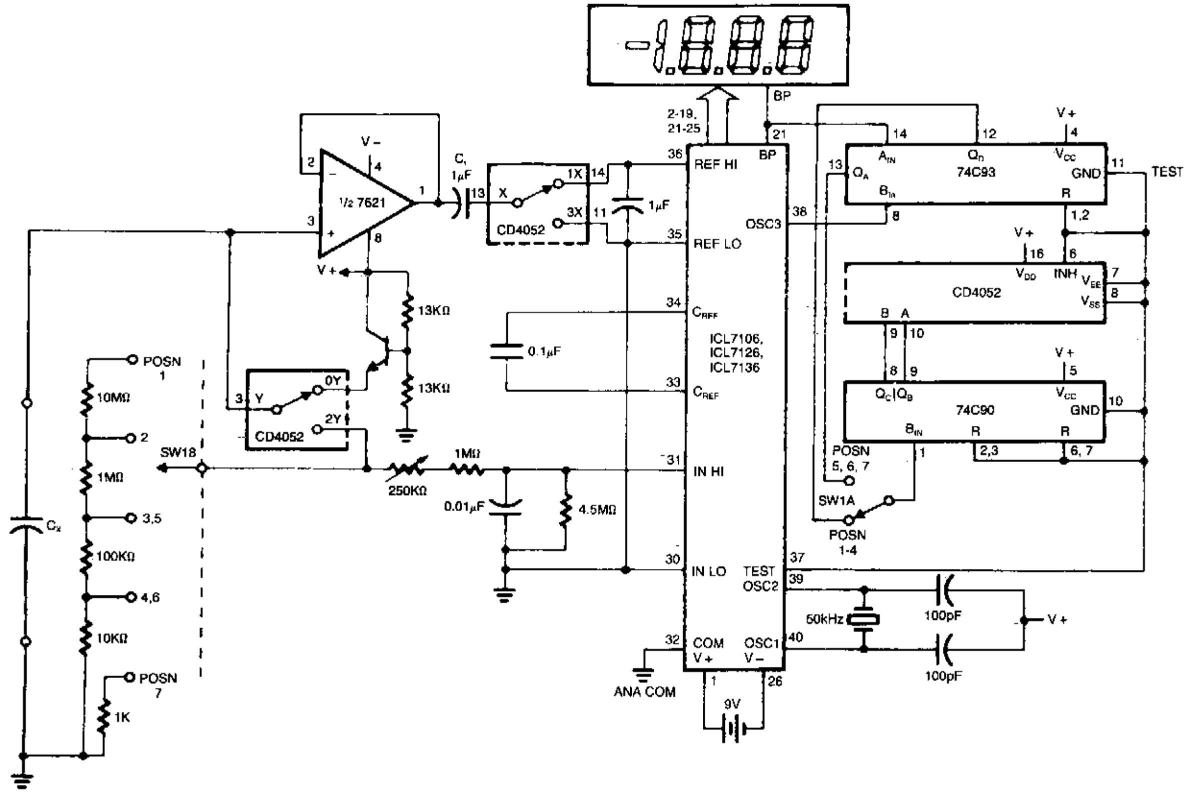
Most European repeaters must be brought up with a 1750-Hz tone. The SK3733 (also known as an ECG1197) IC contains a crystal oscillator and is divided by −256, 1024, 2048, and 4096. A 7168-kHz crystal is used; the divide-by-4096 output produces a 1750-Hz signal.

13

Capacitance Meters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

$3^1/_2$-DIGIT A/D CAPACITANCE METER



	Switch 1A	Switch 1B	Max C	Counter/Switch Phases		
1	10MΩ	6KHz	200pF	0Y	00	Charge C_X
2	1MΩ	6KHz	2nF	1Y	01	ΔV_{CX} on C_{REF}
3	100KΩ	6KHz	20nF	2Y	10	Discharge C_X thru R_{net}
4	10KΩ	6KHz	0.2μF	3X	11	Reset C_1 to Zero
5	100KΩ	60Hz	2μF			
6	10KΩ	60Hz	20μF			
7	1KΩ	60Hz	200μF			

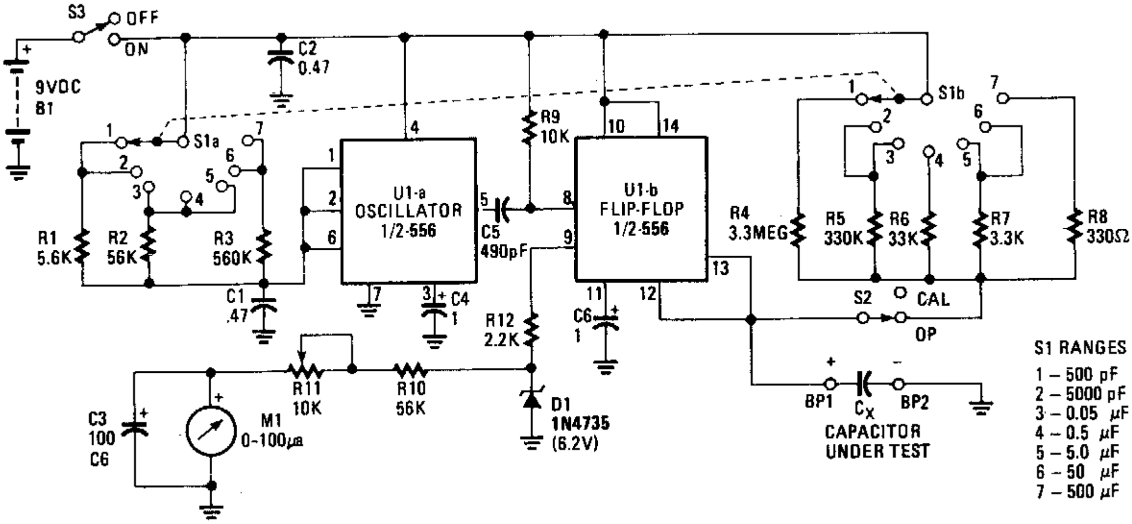
INTERSIL

Fig. 13-1

The circuit charges and discharges a capacitor at a crystal-controlled rate, and stores on a sample-and-difference amplifier the change in voltage achieved. The current that flows during the discharge cycle is averaged, and ratiometrically measured in the a/d using the voltage change as a reference. Range switching is done by changing the cycle rate and current metering resistor. The cycle rate is synchronized with the conversion rate of the a/d by using the externally divided internal oscillator and the internally divided back plane signals. For convenience in timing, the switching cycle takes 5 counter states, although only four switch configurations are used. Capacitances up to 200 μF can be measured, and the resolution on the lowest range is down to 0.1 pF.

The zero integrator time can be set initially at $^1/_3$ to $^1/_2$, the minimum auto-zero time, but if an optimum adjustment is required, look at the comparator output with a scope under worst-case overload conditions. The output of the delay timer should stay low until after the comparator has come off the rail, and is in the linear region (usually fairly noisy).

CAPACITANCE METER



HANDS-ON ELECTRONICS

Fig. 13-2

U1a is an oscillator and U1b the measurement part of the circuit. It converts unknown capacity into a pulse-width modulated signal the same way an automotive dwell meter works. The meter is linear so the fraction or percentage of time that the output is high is directly proportional to the unknown capacitance (*CX* in the schematic). Meter M1 reads the average voltage of those pulses since its mechanical frequency response is low compared to the oscillator frequency of U1a.

14

Carrier-Current Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

CARRIER-CURRENT AUDIO TRANSMITTER

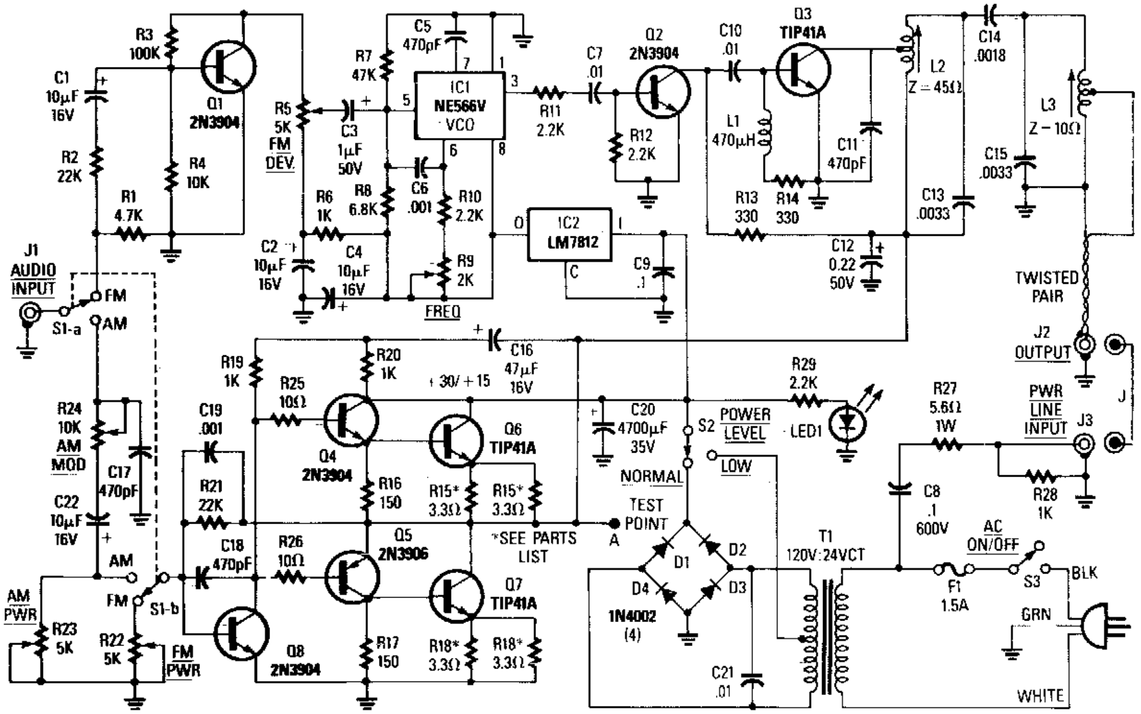


Fig. 14-1

The decision to use either AM, narrowband FM (less than 15 kHz), or wideband FM (greater than 30 kHz) depends on the application. For the transmission of music, FM is better because it has greater noise immunity. For speech or other noncritical applications, AM may be satisfactory. Our transmitter permits either mode by switch selection.

Audio is fed from S1a to either the FM or AM circuitry. Starting with the FM section, amplifier Q1 accepts an audio signal in the 10 Hz to 20 kHz range of about 0.5 V peak-to-peak. The audio gain is adjusted via R5 to provide up to 60 kHz deviation of voltage-controlled oscillator IC1 which is set to nominally 280 kHz. IC1 and Q1 are supplied with a regulated 12 V from IC2. A square-wave signal from IC1 pin 3 drives Q2, and Q2 drives the output amplifier Q3. A coupling network is used to match the nominal 45-Ω output impedance of Q3 to the 10-Ω ac line impedance.

In the AM mode, audio is coupled to Q8 via R24 and then amplified again by transistors Q4 to Q7. The normally stable dc voltage at test point A is thereby varied at an audio rate. Because Q2 and Q3 obtain their dc V_{CC} from test point A, the VCO carrier input to Q2 is amplitude modulated by the varying V_{CC} amplitude. That produces an amplitude-modulated output from the transmitter. Careful setting of carrier level R23 and audio level R24 provides up to 100% modulation. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

CARRIER-CURRENT FM RECEIVER

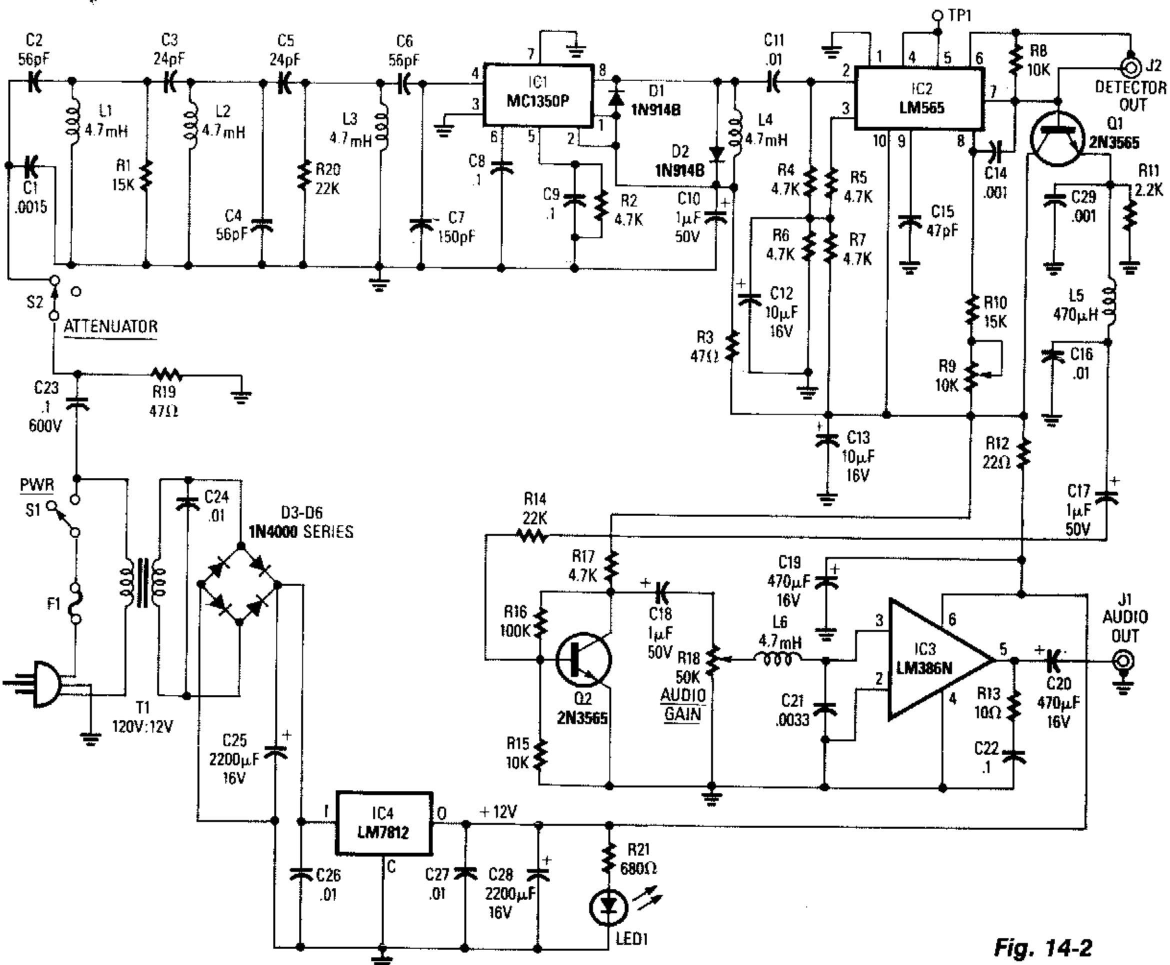


Fig. 14-2

Input signals from the power line are coupled through C23 and R19 to the input filter network. C23 must be rated at 600 volts. Switch S2 is used as an attenuator. Components C2 through C7, L1 through L3, R1, and R20 form a triple-tuned bandpass filter having a passband from 220 – 340 kHz. Signals from the filter are fed to an MC1350P gain block IC, which is used as a tuned rf amplifier.

IC2, the LM565 PLL, is used as an FM demodulator. Pins 8 and 9 are connected to an internal VCO and components R9, R10, and C15 set the VCO's free running frequency. The VCO signal and the input signal from pin 2 are compared in the phase detector. The output from the phase detector is internally amplified, and then appears at pin 7. The output at pin 7 is a replica of the original modulation on the FM input signal to the receiver; the output at pin 7 is therefore the recovered audio. C17 and R14 couple audio to the base of Q2, which, in conjunction with R15, R16, R17, and C18, form an audio amplifier that brings the recovered audio up to around 1 V peak-to-peak. The signal is then fed into an LM386N audio amplifier, which can deliver up to 1/2 W of audio, coupled via C20, to any standard 8-Ω external speaker. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

CARRIER-CURRENT AM RECEIVER

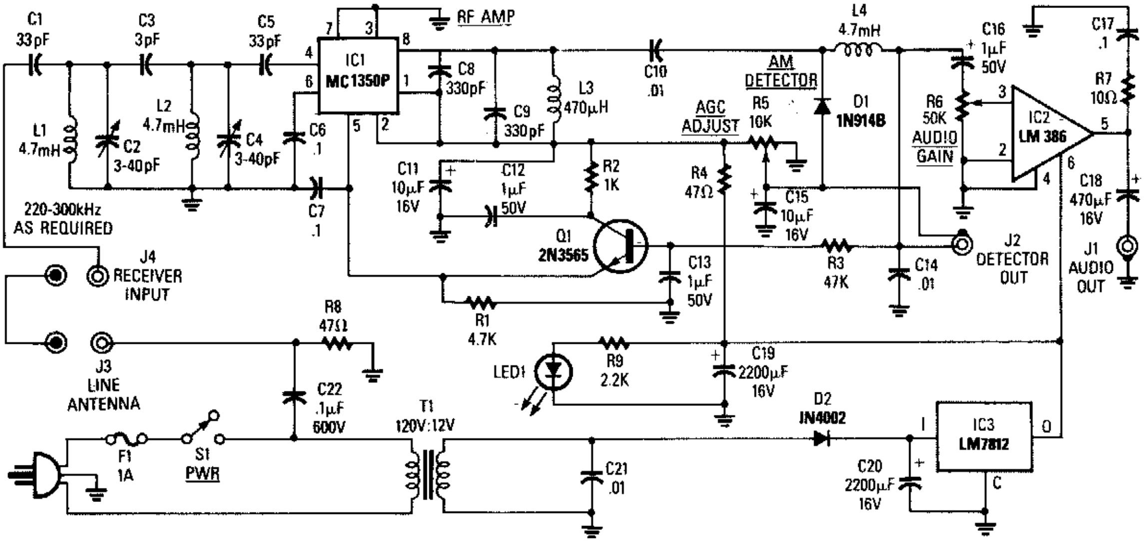


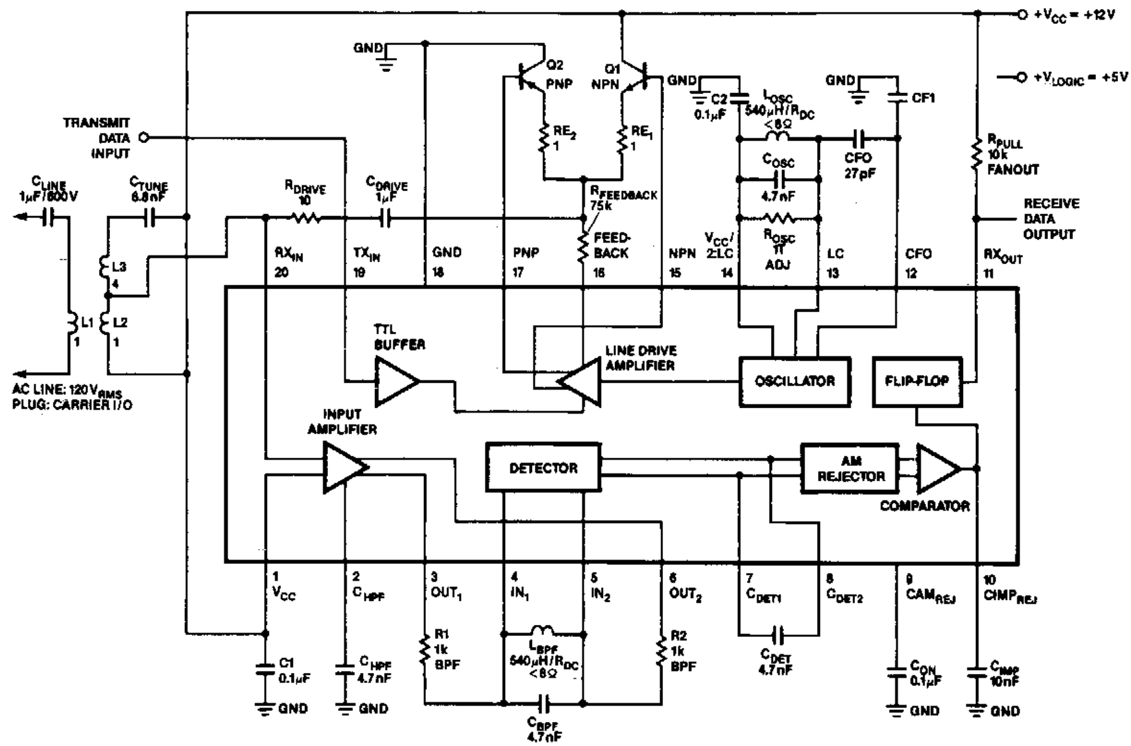
Fig. 14-3

The AM Tuned Radio Frequency (TRF) receiver, has a sensitivity of about 1 mV at the input for an audio output of 1/2 W. Capacitor C22 couples audio signals from the power line to the PC board—it must be rated at 600 Vdc. R8 will cause F1 to blow, if C22 shorts. The signal from C22 goes to a tuned network (C1 through C5, L1, and L2) that has a 20-kHz bandwidth, which allows only the desired signal to pass through.

IC1 is a *gain block* i-f chip that has AGC capability and approximately 60 dB of gain. Components C8, C9, and L3, which are placed across the output of IC1, are broadly resonant around 280 kHz. C10 couples rf to detector-diode D1, which is used as an envelope detector.

The detector output is taken from C14, which sets the upper frequency limit at about 10 kHz or so. By reducing the value of C14, high frequency response can be obtained. The detector output is connected to an external jack. Audio components are fed to audio-gain control R6, through C16 to IC2, an audio amplifier. C18 couples up to 1/2 watt of audio to an external speaker. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

POWER-LINE MODEM



SIGNETICS

Fig. 14-4

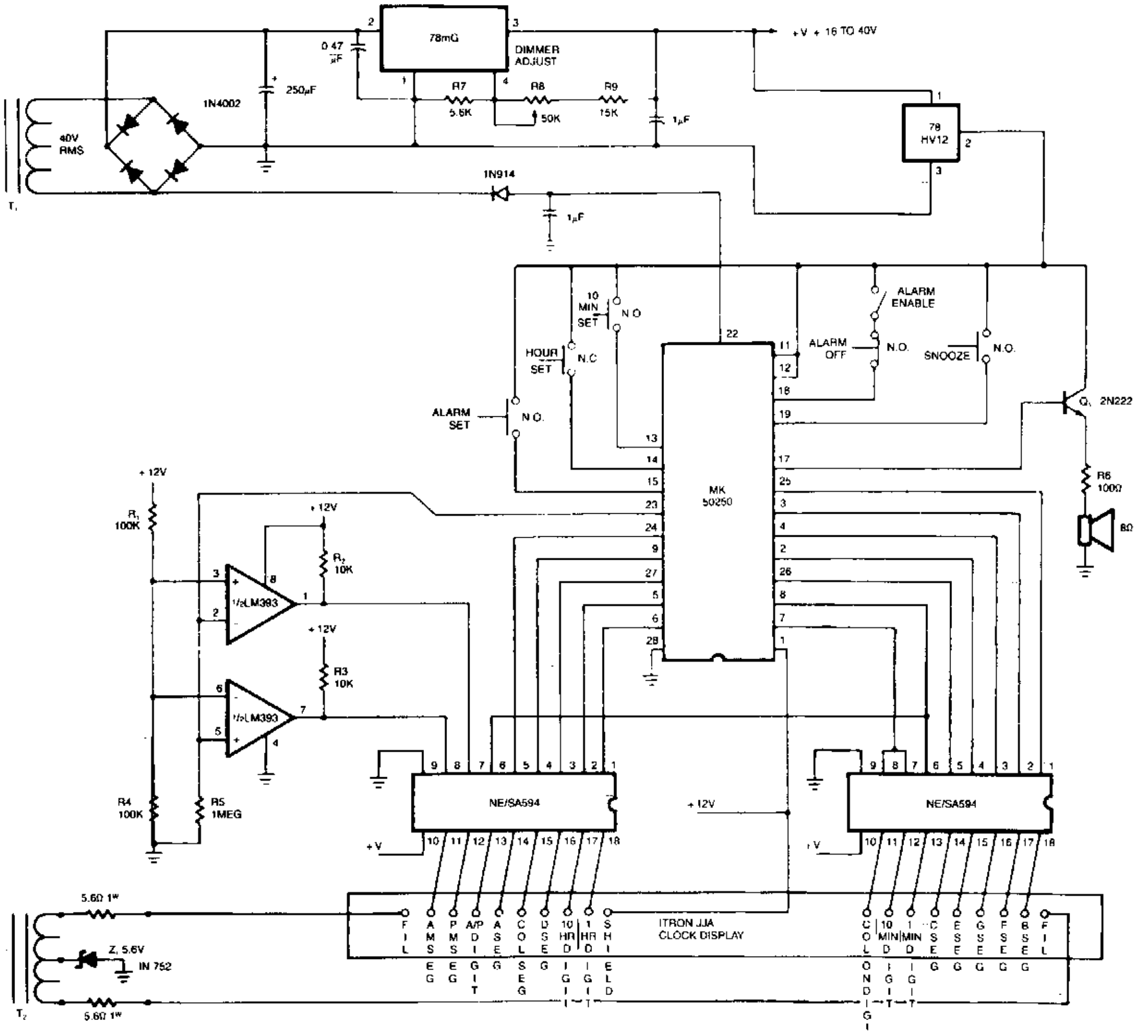
In the 100-kHz application from left to right, the coupling network feeds into the receiver section on the bottom of the chip. (The external components are summarized later.) The receive data output is pulled up via R_{PULL} = 10 KΩ. A minimum current of 10 mA sets the voltage drop across R_{PULL}. Another voltage supply, V_{LOGIC}, is shown if the user wants to have the output sent at TTL levels. Across the top is the transmitter section; going from right to left, the oscillator network, the class AB output stage (note feedback resistor $R_{FEEDBACK}$) and the drive section. The LC values on the oscillator network should match those on the bandpass filter in the receiver. The drive stage feeds into the coupling network and back into the receive section. This enables the on-chip collision detection with listen-while-talking capability. This effect can be cancelled, although the transmitter will still be connected to the receiver.

15

Clock Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DIGITAL CLOCK WITH ALARM



SIGNETICS

Fig. 15-1

OSCILLATOR/CLOCK GENERATOR

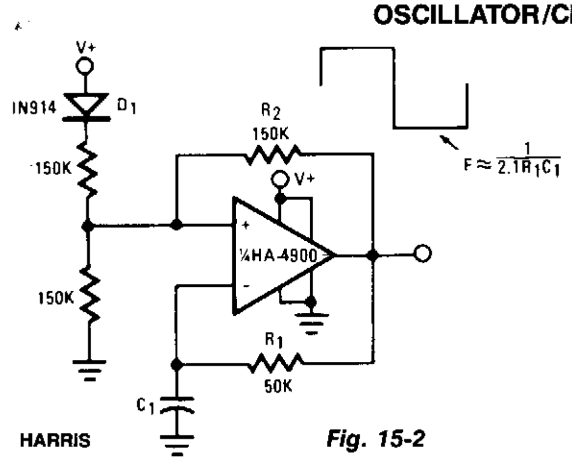


Fig. 15-2

This self-starting fixed-frequency oscillator circuit gives excellent frequency stability. R1 and C1 comprise the frequency-determining network, while R2 provides the regenerative feedback. Diode D1 enhances the stability by compensating for the difference between V_{OH} and V_{SUPPLY}. In applications where a precision clock generator up to 100 kHz is required, such as in automatic test equipment, C1 might be replaced by a crystal.

SINGLE OP AMP CLOCK

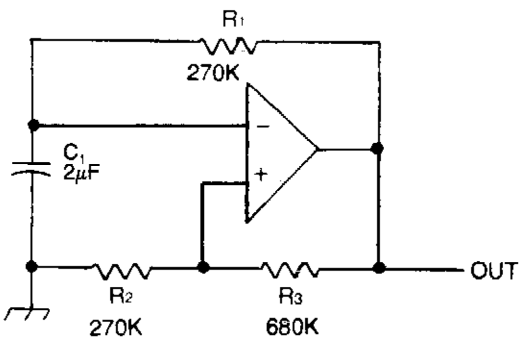


Fig. 15-3

Capacitor C1 is charged through timing resistor R1 when the clock output is high. When C1 reaches the upper threshold voltage, the output signal decreases, and then C1 discharges through R1 until its voltage reaches the lower threshold point. When this happens, the output increases again and the cycle repeats itself. Using the parts values shown results in a frequency of 1 Hz. The output frequency can be adjusted by trimming the value of R1 slightly.

WIDE-FREQUENCY TTL CLOCK

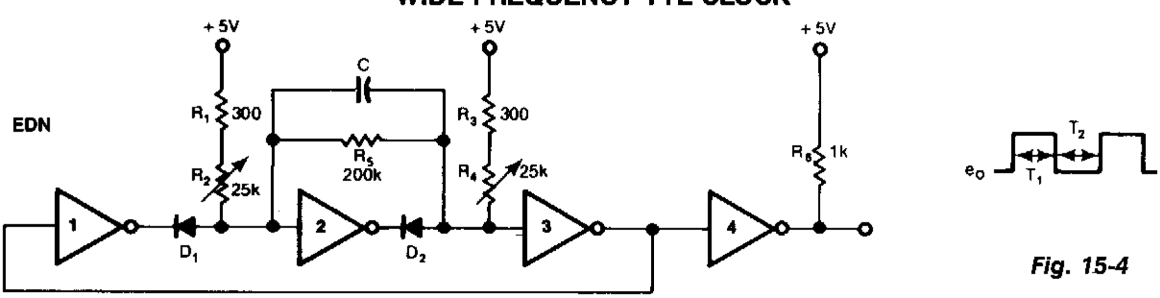


Fig. 15-4

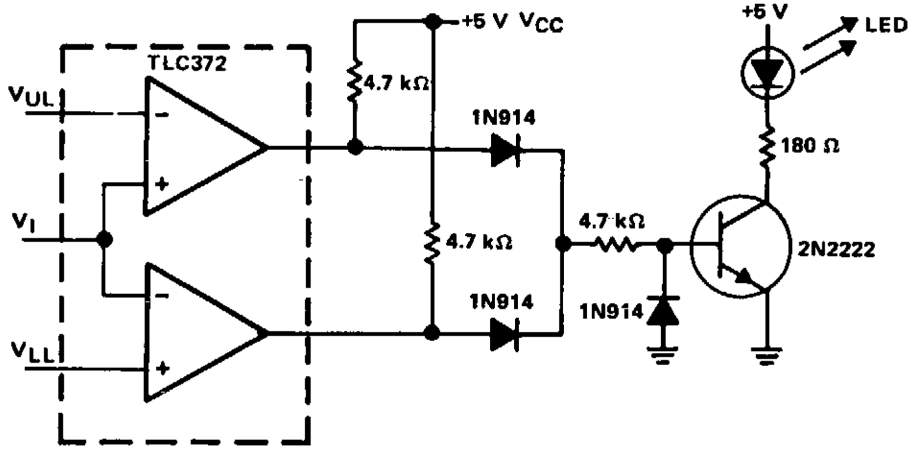
This free-running TTL square-wave oscillator has a variable frequency output over a 20:1 range or better through use of four of the six inverters in an SN7404 chip and the additional components shown. Frequency of oscillation is determined by the capacitor and the settings of potentiometers R2 and R4; the first pot controls width T1 and the second controls width T2 of the square-wave output. These adjustments are not completely independent.

16

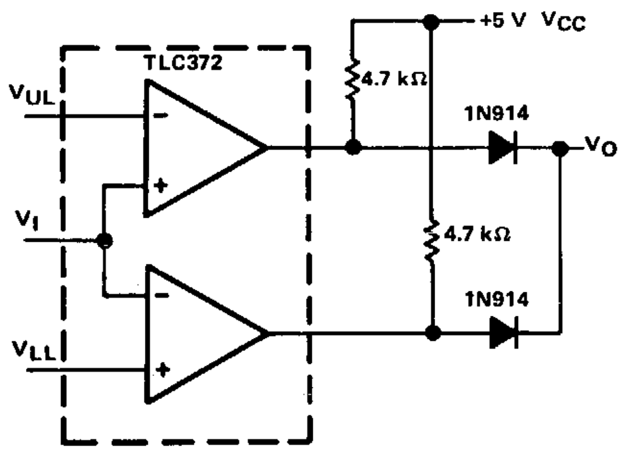
Comparators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

WINDOW COMPARATOR



Window Comparator with LED Indicator



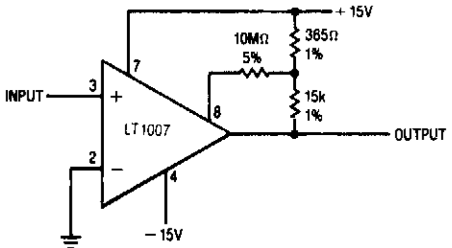
Reprinted by permission of Texas Instruments.

Basic Window Comparator

Fig. 16-1

A window detector is a specialized comparator circuit designed to detect the presence of a voltage between two prescribed limits; that is, within a voltage *window*. This circuit is implemented by logically combining the outputs of two single-ended comparators by the IN914 diodes. When the input voltage is between the upper limit, V_{UL}, and the lower limit, V_{LL}, the output voltage is zero; otherwise it equals a logic high level. The output of this circuit can be used to drive a logic gate, LED driver, or relay driver circuit. The circuit shown in Fig. 16-1 shows a 2N2222 npn transistor being driven by the window comparator. When the input voltage to the window comparator is outside the range set by the V_{UL} and V_{LL} inputs, the output changes to positive, which turns on the transistor and lights the LED indicator.

MICROVOLT COMPARATOR WITH HYSTERESIS

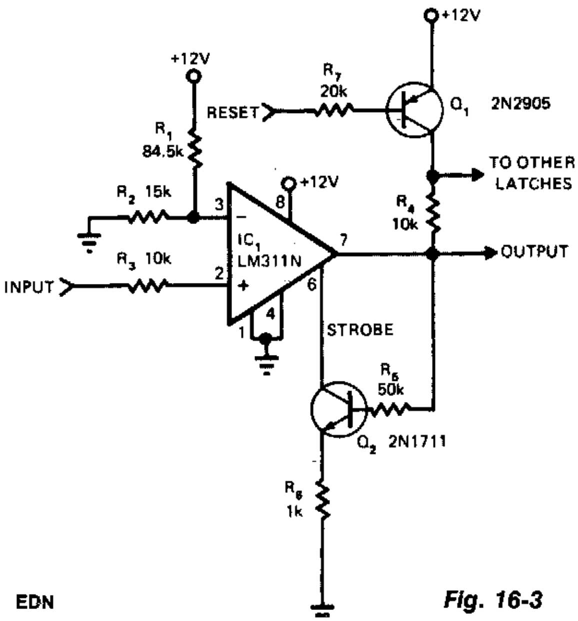


LINEAR TECHNOLOGY

Fig. 16-2

Positive feedback to one of the nulling terminals creates approximately 5 μV of hysteresis. The output can sink 16 mA; the input offset voltage is typically changed less than 5 μV because of the feedback.

COMPARATOR/LATCH

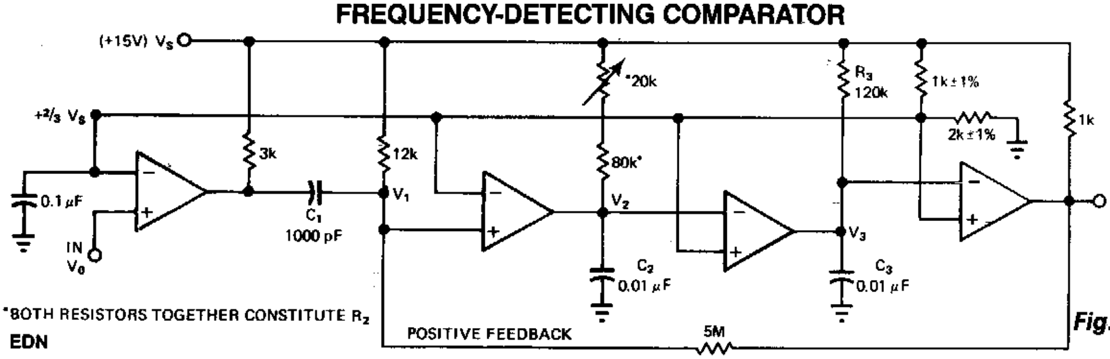


EDN

Fig. 16-3

The primary advantage of this circuit, when compared to other comparators, is its ability to latch after the input has reached a predetermined threshold level. When the input exceeds the threshold level, the LM311N output increases. This transition enables the strobe input, preventing the output from falling low. A high-level voltage on the reset input will turn off Q1, thereby removing the supply voltage from the open collector output of the LM311N. With no supply to the strobe input, the latch condition is removed and the output is again allowed to follow the input excursions. The LM311N will operate with a wide variety of supply voltage levels, ranging from dual ±15 V to a single 5 V level that provides compatibility with digital IC logic. If more than one latch is used with a common reset, all the pull-up resistors may be connected to Q1's collector.

FREQUENCY-DETECTING COMPARATOR



*BOTH RESISTORS TOGETHER CONSTITUTE R_2

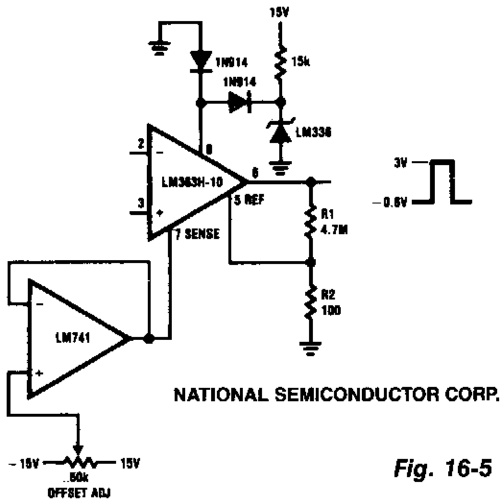
EDN

Fig. 16-4

FREQUENCY-DETECTING COMPARATOR (*CONT.*)

A quad comparator forms the basis of a frequency detector that is faster and less expensive than more complex versions designed around frequency-to-voltage converter chips. Positive feedback through a 5-MΩ resistor allows the circuit to resolve changes as small as two percent; the output responds to those changes in about one cycle. When the input frequency is high, V_2 is pulled low; it's never allowed to exceed 2/3 V. When the input frequency is lower than the limit, V_2 exceeds 2/3 V once each cycle, but V_3 is held below that limit. The trip frequency is defined by $F = 1/(1.1R_2C_2)$. R2 can be adjusted to permit trimming of the trip point, but the value of R3 must remain larger than R2.

PRECISION COMPARATOR WITH BALANCED INPUTS AND VARIABLE OFFSET



$$t_{PD} \approx 15\ \mu s \text{ at 1-mV overdrive}$$

$$V_{OUT} = V_2 + 0.6\ V$$

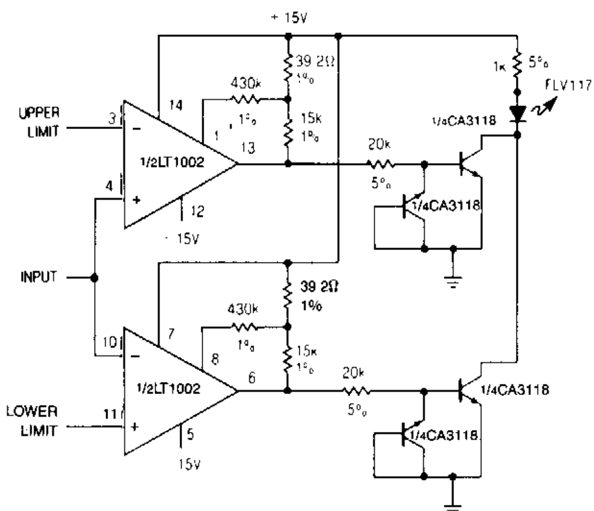
$$\text{Hysteresis} = \frac{V_{OUT}}{R\,(R_1 + R_2)} = 2\ mV$$

$$\text{Offset} = V_{SENSE}/G$$
$$\pm\ 1.3 \text{ range}$$

NATIONAL SEMICONDUCTOR CORP.

Fig. 16-5

DUAL LIMIT MICROVOLT COMPARATOR



When the upper or lower limit is exceeded, the LED lights up. Positive feedback to one of the nulling terminals creates 5 to 20 μV of hysteresis on both amplifiers. This feedback changes to offset voltage of the LT1002 by less than 5 μV. Therefore, the basic accuracy of the comparator is limited only by the low offset voltage of the LT1002.

LINEAR TECHNOLOGY

Fig. 16-6

WINDOW COMPARATOR

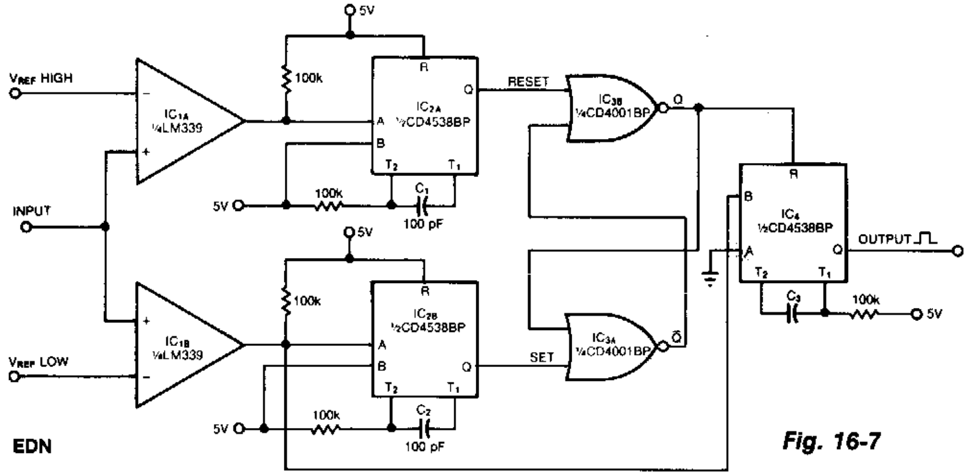


Fig. 16-7

This window comparator generates an output pulse for each event that occurs within a specified window. That is, each output pulse signifies an input voltage pulse or level change that exceeds V_{REFLOW}, but not $V_{REFHIGH}$. The monostable multivibrators, IC2A and IC2B, produce a 10-μs pulse at their Q output in response to a rising edge at their A input. Comparator IC1B produces a rising edge when the input exceeds V_{REFLOW}, and comparator IC2A produces a rising edge when the input exceeds $V_{REFHIGH}$. The NOR gates, IC3A and IC3B, form a bistable latch whose Q output, when low, disables IC4. IC4, unless disabled, produces output pulses in response to falling edges at the IC1B comparator output. You set the width of these pulses by selecting the value of C3. The circuit can handle an input waveform containing 0 to 2 V amplitudes and 10-Hz to 10-kHz frequency components.

FOUR-CHANNEL COMPARATOR

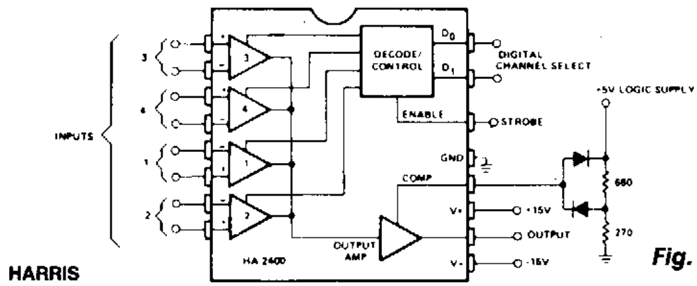


Fig. 16-8

When operated as an open loop without compensation, the HA-2400 becomes a comparator with four selectable input channels. The clamping network at the compensation pin limits the output voltage to allow DTL or TTL digital circuits to be driven with a fanout of up to ten loads.

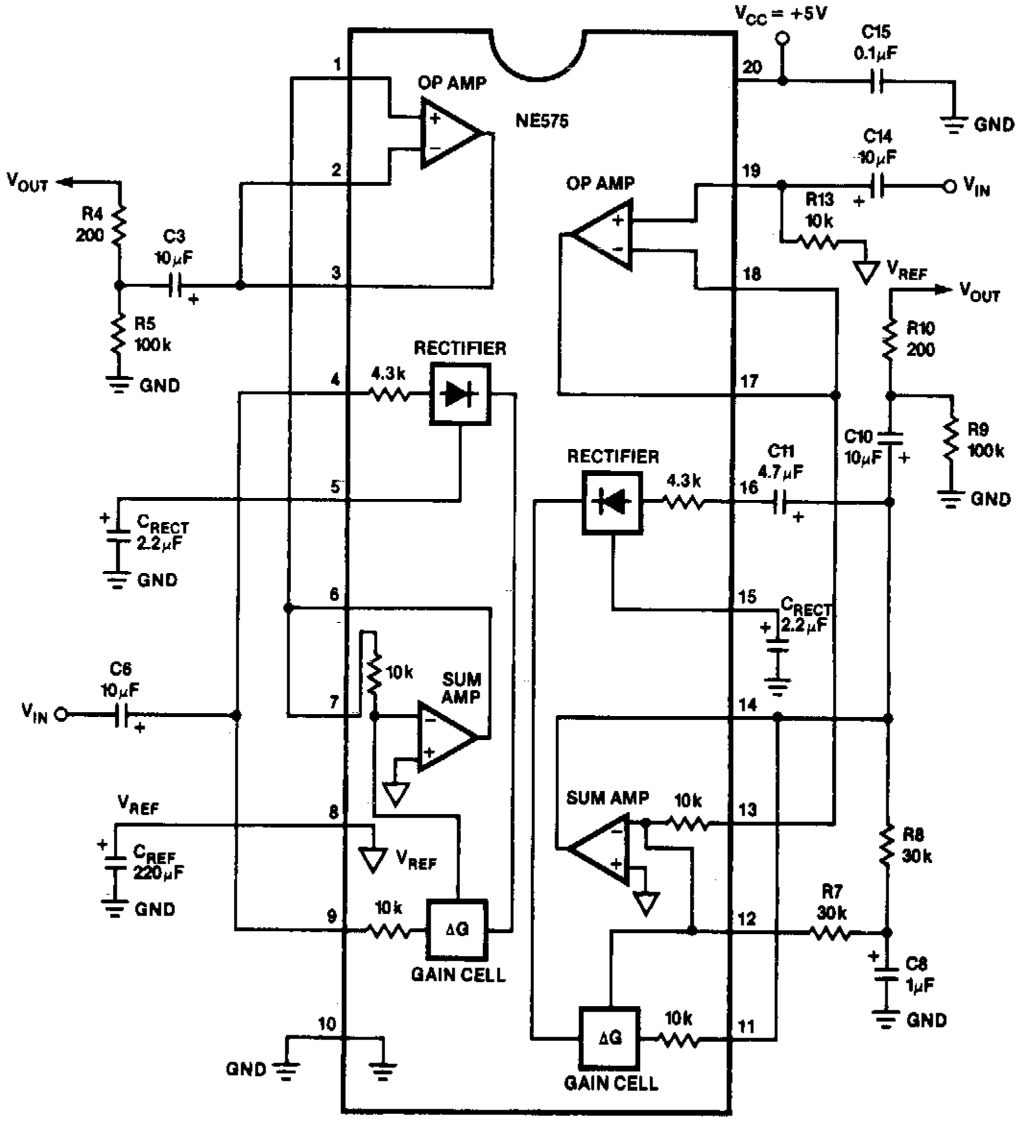
The circuit can be used to compare several signals against each other or against fixed references; or a single signal can be compared against several references. A window comparator, which assures that a signal is within a voltage range, can be formed by monitoring the output polarity, while rapidly switching between two channels with different reference inputs and the same signal input.

17

Compressor/Expander Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

LOW-VOLTAGE COMPANDER



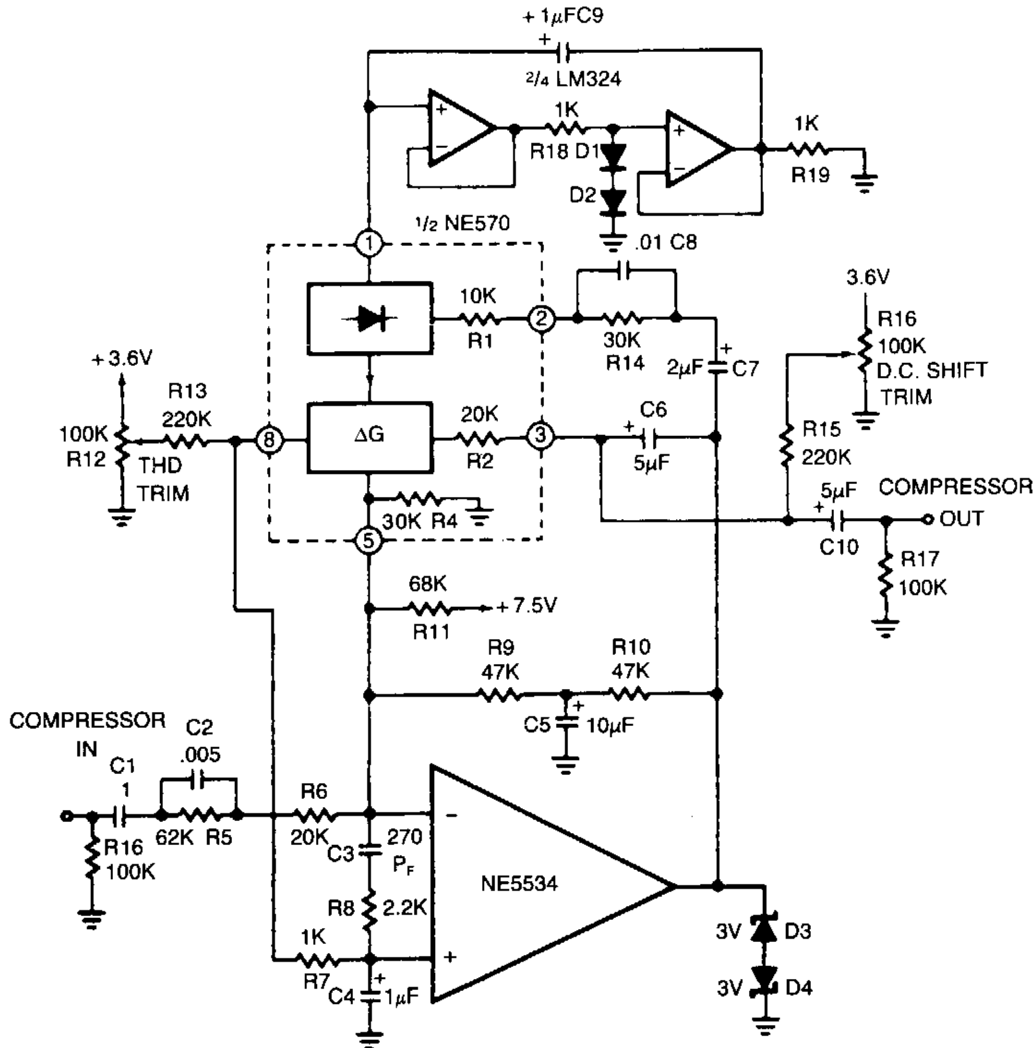
NOTE:
Left channel in expander mode; right channel in compressor mode.
For additional information, call the factory.

SIGNETICS

Fig. 17-1

The NE575 is a dual-gain control circuit designed for low voltage applications. The NE575's channel 1 is an expander, while channel 2 can be configured either for expander, compressor, or automatic level controller (ALC) applications.

HI-FI COMPRESSOR WITH PRE-EMPHASIS

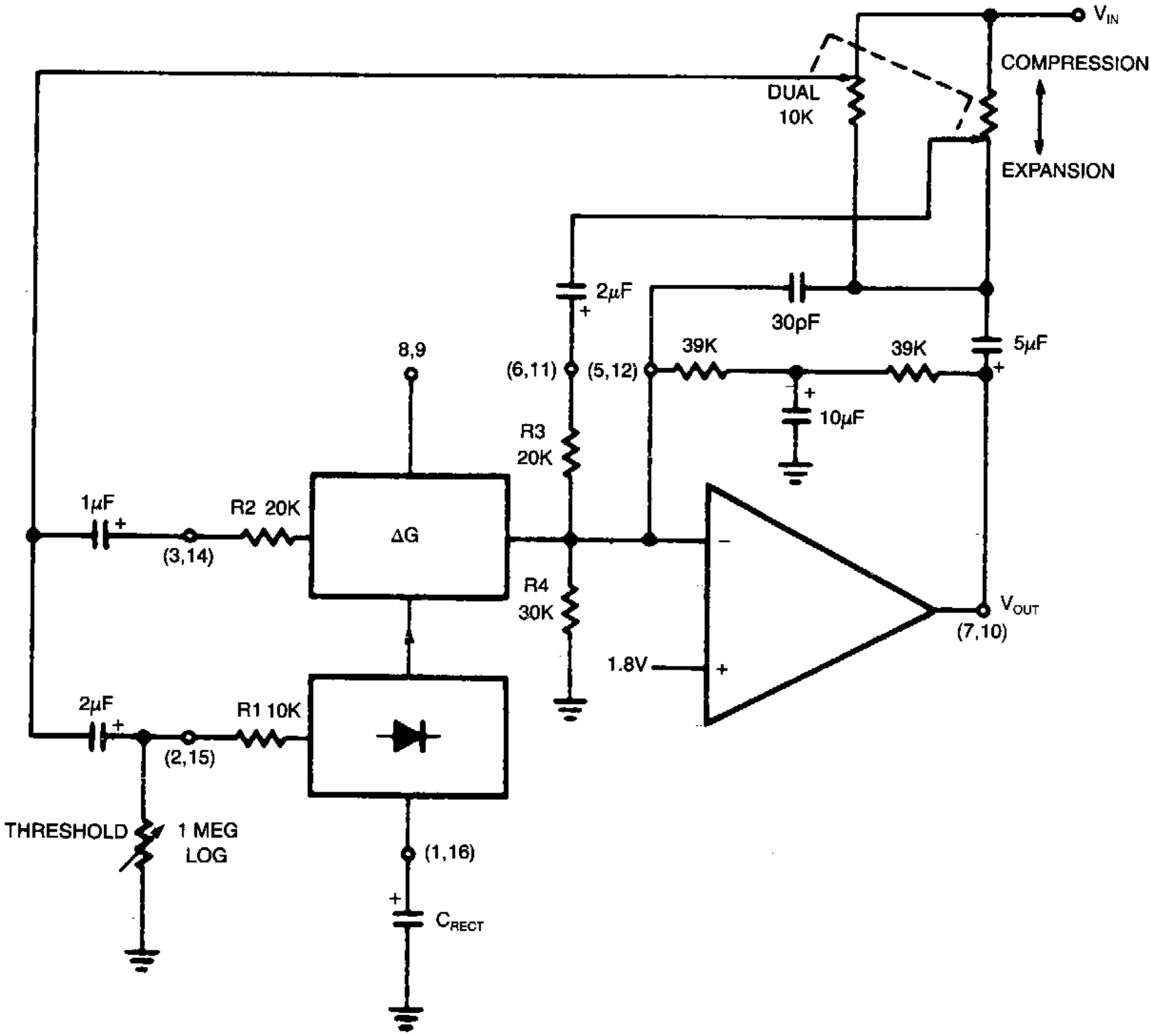


SIGNETICS

Fig. 17-2

The compressor contains a high-frequency, pre-emphasis circuit (C2, R5, and C8, R14), which helps solve this problem. Matching de-emphasis on the expander is required. More complex designs could make the pre-emphasis variable.

VARIABLE SLOPE COMPRESSOR/EXPANDER

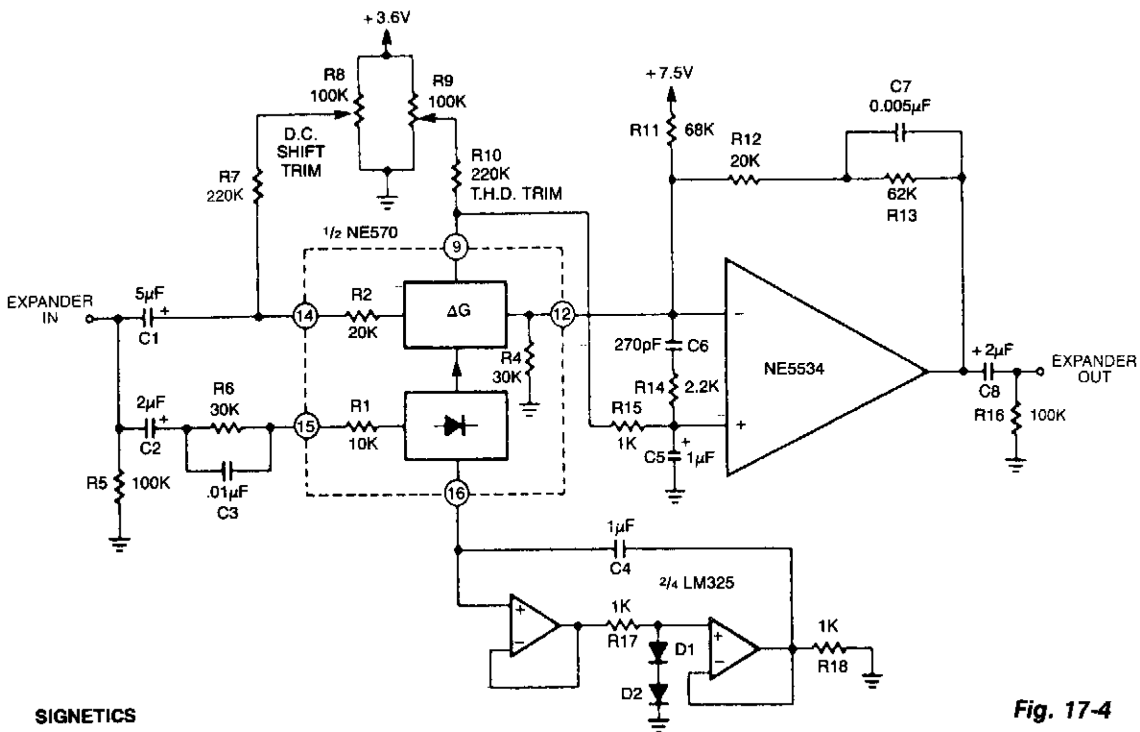


SIGNETICS

Fig. 17-3

Compression and expansion ratios other than 2:1 can be achieved by the circuit shown. Rotation of the dual potentiometer causes the circuit hook-up to change from a basic compressor to a basic expander. In the center of rotation, the circuit is 1:1, has neither compression nor expansion. The (input) output transfer characteristic is thus continuously variable from 2:1 compression to 1:2 expansion. If a fixed compression or expansion ratio is desired, proper selection of fixed resistors can be used instead of the potentiometer. The optional threshold resistor will make the compression or expansion ratio deviate towards 1:1 at low levels. A wide variety of (input) output characteristics can be created with this circuit.

HI-FI EXPANDER WITH DE-EMPHASIS



SIGNETICS

Fig. 17-4

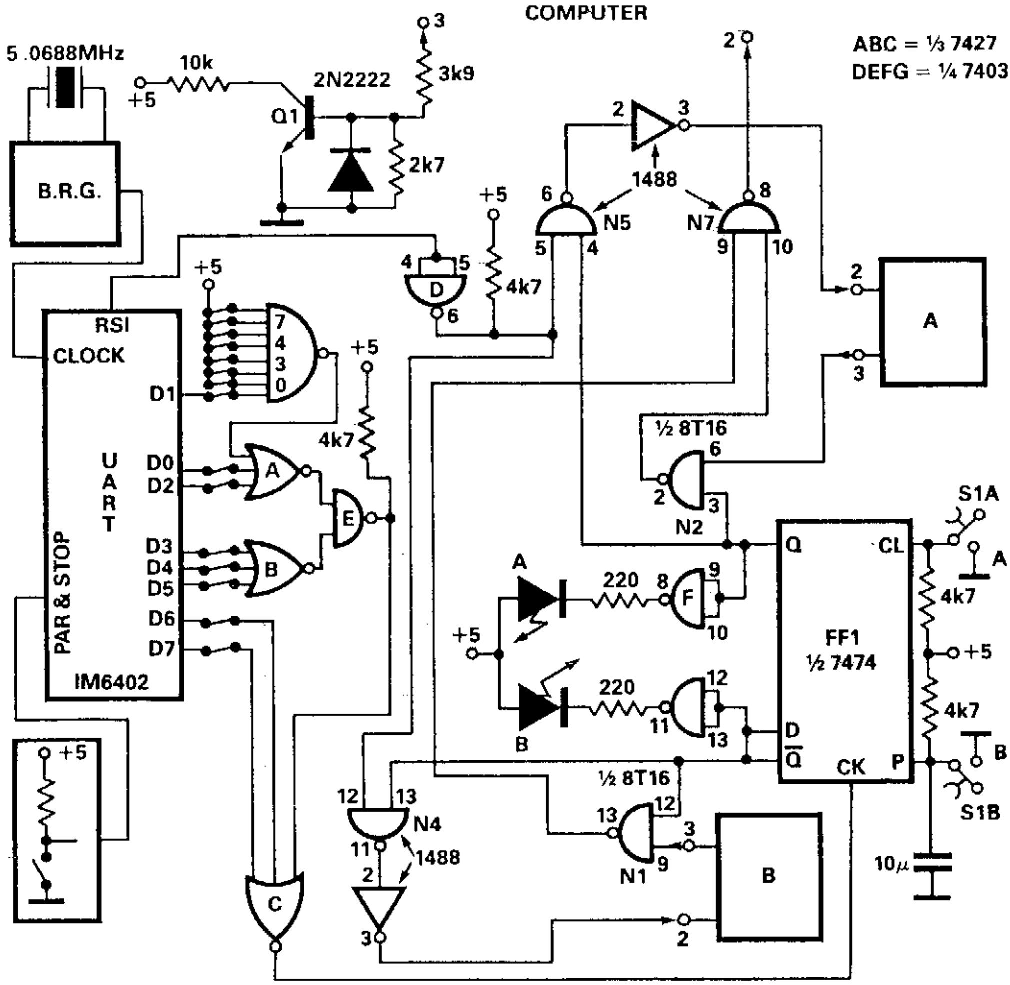
The expander to complement the compressor is shown. An external op amp is used for high slew rate. Both the compressor and expander have unity gain levels of 0 dB. Trim networks are shown for distortion (THD) and dc shift. The distortion trim should be done first, with an input of 0 dB at 10 kHz. The dc shift should be adjusted for minimum envelope bounce with tone bursts. When applied to consumer tape recorders, the subjective performance of this system is excellent.

18

Computer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

AUTOMATIC RS-232 DATASELECTOR



ELECTRONIC ENGINEERING

Fig. 18-1

With this dataselector, only one RS-232 port is used to connect two RS-232 devices (i.e., printer, plotter, etc.) with a mini- or microcomputer. The operation is very simple. Power on will reset FFI (Q_{FFI} = Low), which enables gates N1, N5, and N7. Now communication between computer and device B is possible. Detection of the switch command, i.e., Control B character = CHR$(2), selectable with wire-wrap pins, on the parallel outputs of the UART (IM 6402 or equivalent) will set: Q_{FFI} = High. Gates N2, N5, and N7 are open, so device A is connected with the computer until Control B character is detected again.

Transistor Q1 converts RS-232 levels to TTL levels while two LEDs indicate whether device A or B is linked. The baud-rate generator provides the 16× clock needed for the UART. Any baud rate ranging from 50 to 19200 can be selected. Manual control of the selector is available with toggleswitch S1.

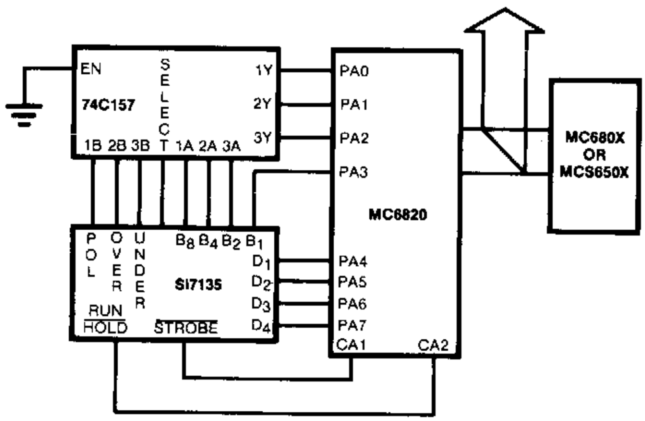


Fig. 18-2a

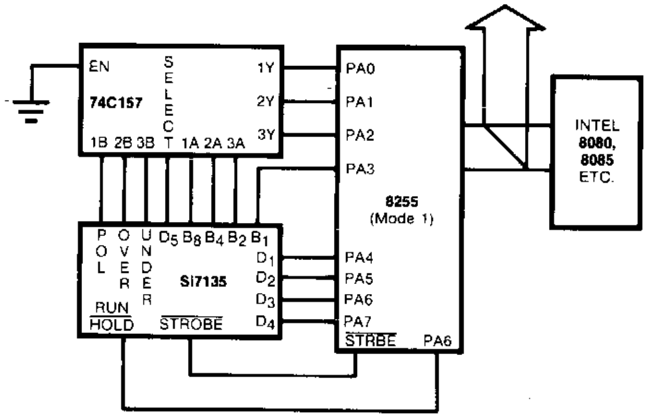


Fig. 18-2b

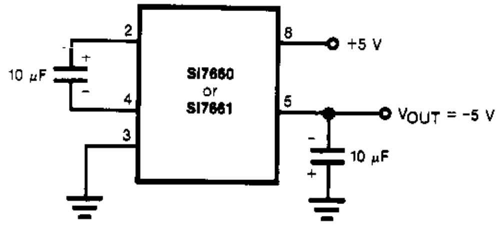


Fig. 18-2c

SILICONIX

Circuits to interface the Si7135 directly with two popular microprocessors are shown in Figs. 18-2a and b. The 8080/8048 and the MC6800 families with 8-bit words need to have polarity, overrange, and underrange multiplexed onto the digit 5 word. In each case, the microprocessor can instruct the ADC when to begin a measurement and when to hold this measurement. The Si7135 is designed to work from ±5 V supplies. However, if a negative supply is not available, it can be generated using 2 capacitors, and an inexpensive Si7660 or Si7661 IC, as shown in Fig. 18-2c.

RGB BLUE BOX

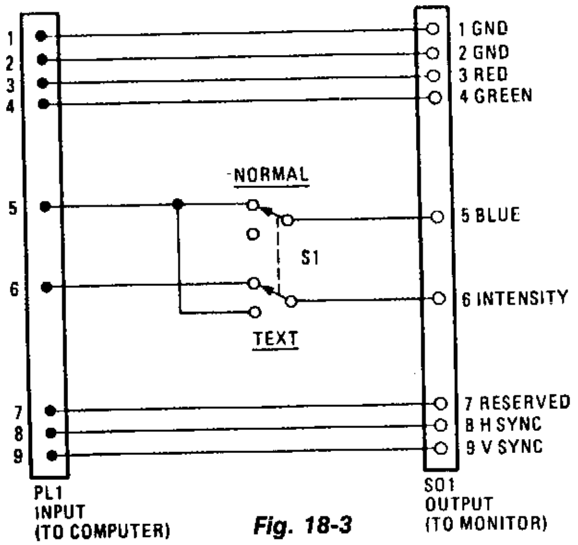
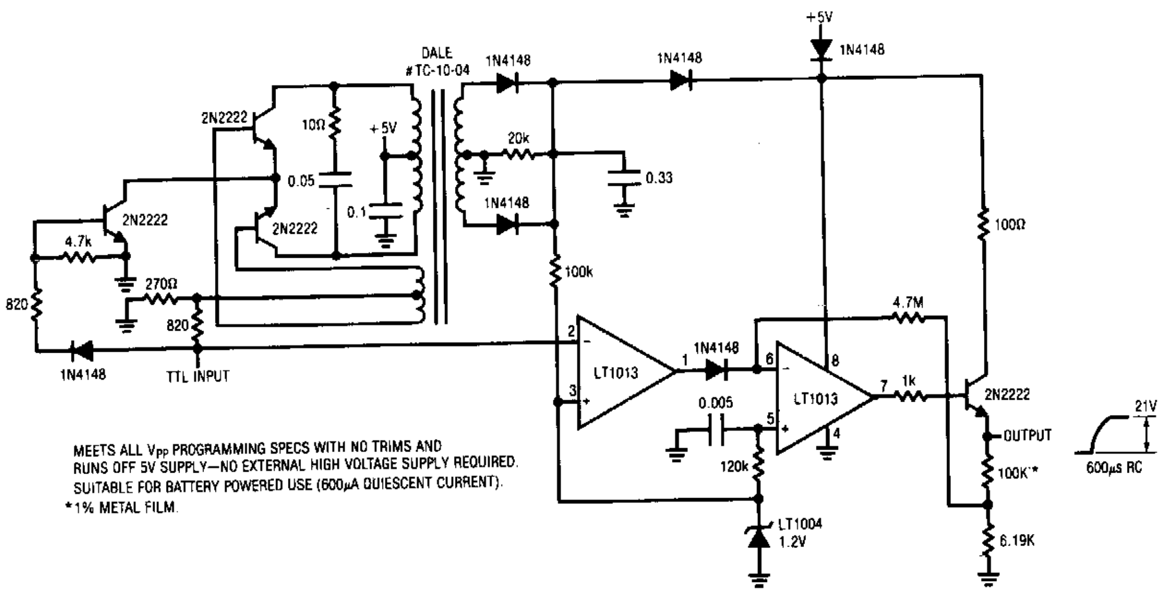


Fig. 18-3

HANDS-ON-ELECTRONICS/POPULAR ELECTRONICS

The RGB blue box turns your PC's RGB-monitor screen blue at the flip of a switch. That is, it makes your computer display bright white text on a blue background, instead of the usual low-intensity white on black. The RGB blue box connects between your IBM PC color graphics adapter, or equivalent, and your RGB color monitor. By flipping a switch, you choose between two modes. One mode passes the signal from the PC to the monitor unaltered; the other transforms it to make text more readable. The monitor has four TTL-level inputs—red, green, blue, and intensity—and it interprets disconnected wires as on. That's why the screen turns white if you disconnect the monitor from the computer, and blue if you disconnect only the blue line. Instead of just discarding the blue signal, the blue box reroutes it to the intensity input. As a result, most of the text colors come out intensified.

5V-POWERED EEPROM PULSE GENERATOR



MEETS ALL V_{PP} PROGRAMMING SPECS WITH NO TRIMS AND RUNS OFF 5V SUPPLY—NO EXTERNAL HIGH VOLTAGE SUPPLY REQUIRED. SUITABLE FOR BATTERY POWERED USE (600μA QUIESCENT CURRENT).
*1% METAL FILM.

LINEAR TECHNOLOGY CORP.

Fig. 18-4

ONE-OF-EIGHT CHANNEL TRANSMISSION SYSTEM

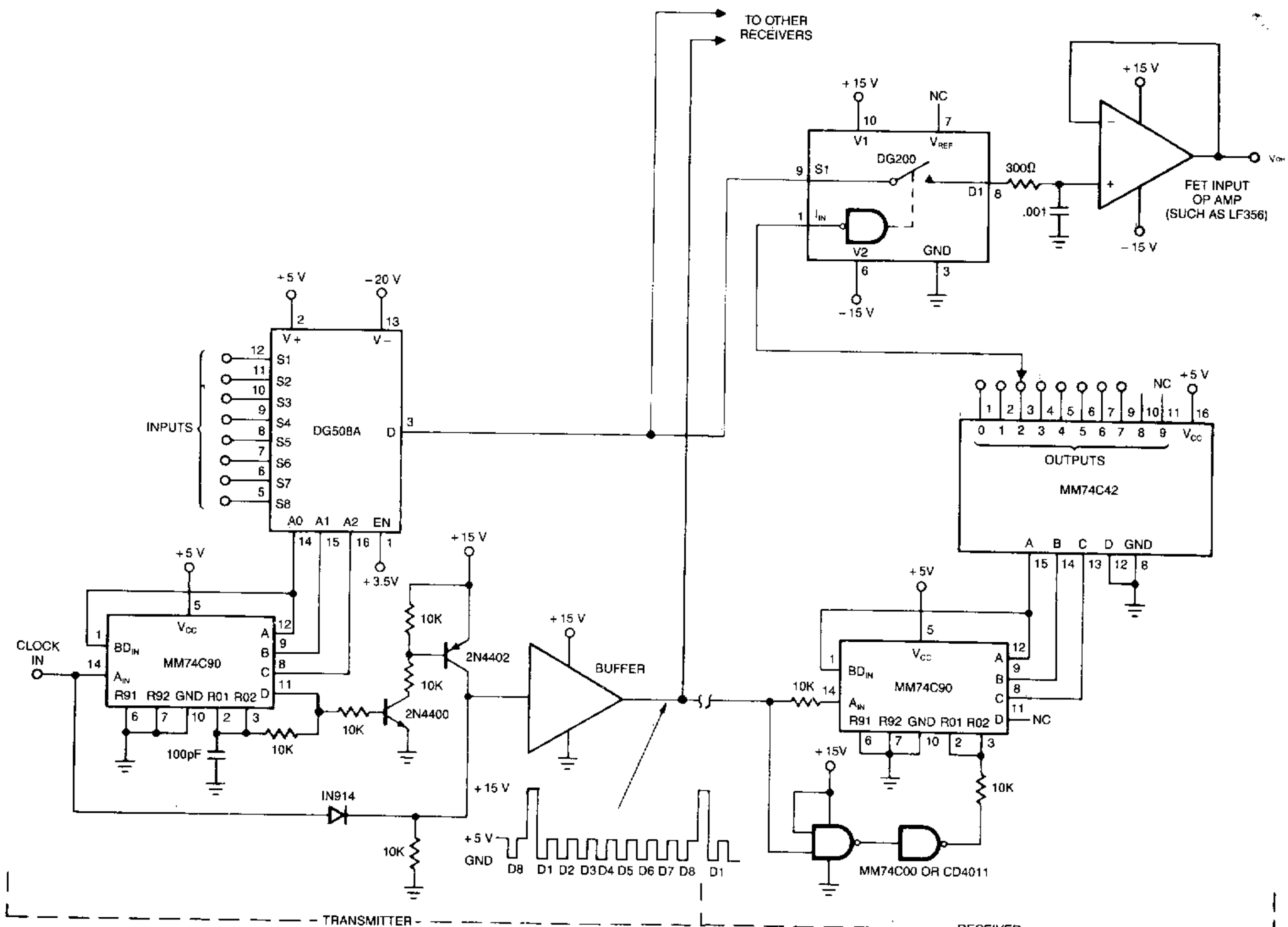
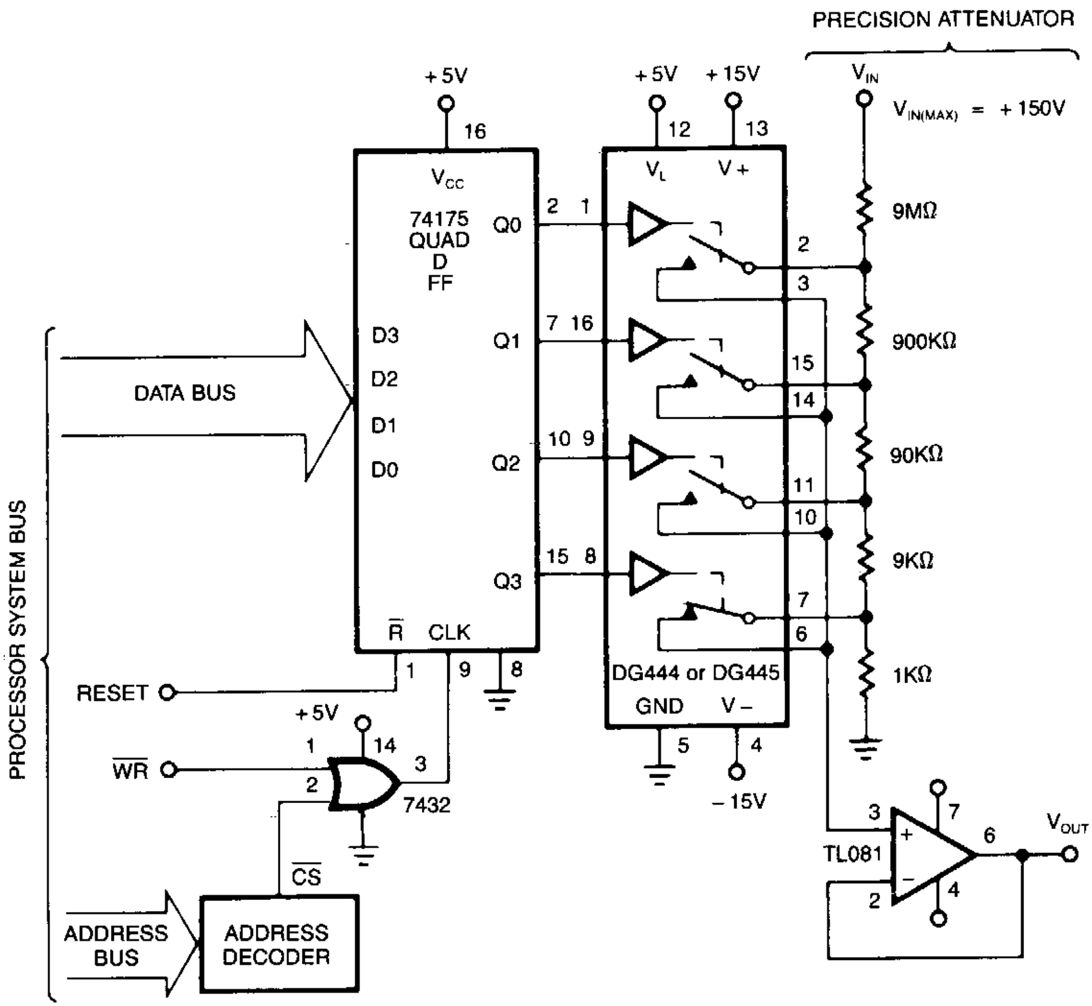


Fig. 18-5

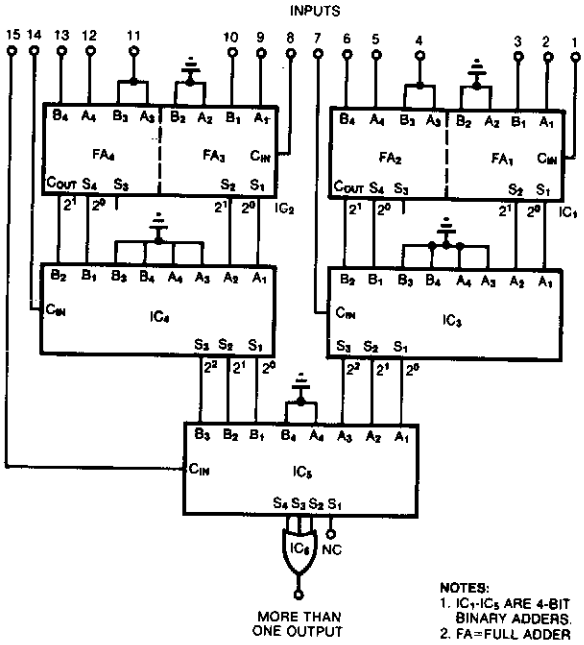
MICROPROCESSOR-CONTROLLED ANALOG SIGNAL ATTENUATOR



SILICONIX

Fig. 18-6

MULTIPLE INPUT DETECTOR



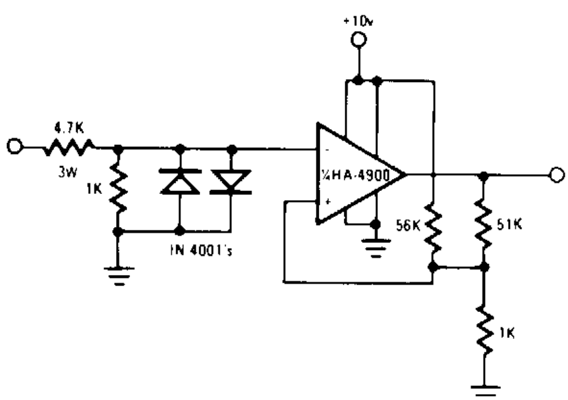
This circuit determines whether more than one input in a group of digital inputs is active. It provides a digital measure of the number of active inputs, and it allows you to establish a threshold for majority-decision applications. That is, whether the number of active inputs is more than, less than, or equal to a value between 1 and 15. You can monitor more inputs by cascading the adders.

Each binary adder, IC1 and IC2, forms two full adders (FAs). Each FA monitors three input lines and generates a 2-bit output representing the number of inputs active. IC3 and IC4, by summing the outputs of two FAs plus an input line, individually measure how many in a group of seven inputs are active. Similarly, by monitoring the 3-bit outputs of IC3 and IC4 plus one input, IC5 measures how many in the group of 15 are active. The OR gate, IC6, simply indicates whether more than one input is active.

EDN

Fig. 18-7

RS-232-TO-CMOS LINE RECEIVER



This RS-232 type line receiver to drive CMOS logic uses a Schmitt-trigger feedback network to give about 1-V input hysteresis for added noise immunity. A possible problem in an interface which connects two pieces of equipment, each plugged into a different ac receptacle, is that the power line voltage might appear at the receiver input when the interface connection is made or broken. The two diodes and a 3-W input resistor will protect the inputs under these conditions.

HARRIS

Fig. 18-8

RS-232C LED CIRCUIT

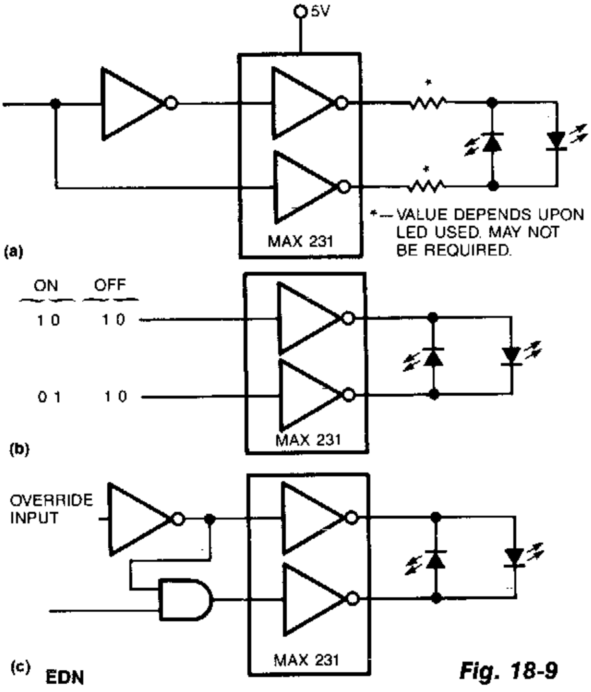


Fig. 18-9

Use a pair of Maxim's 5V-powered MAX231 RS-232C transmitters as drivers to obtain a 2-color LED. The transmitters require only a single-ended, 5-V input to generate ±10 V internally. Their outputs are short-circuit-proof and can supply as much as 10 mA—enough to drive most LEDs. Depending on which LED you select, their current-limiting feature might also eliminate the need for external series resistors. Using the simple circuits, you can implement a variety of functions.

SPARE FLIP-FLOP INVERTER

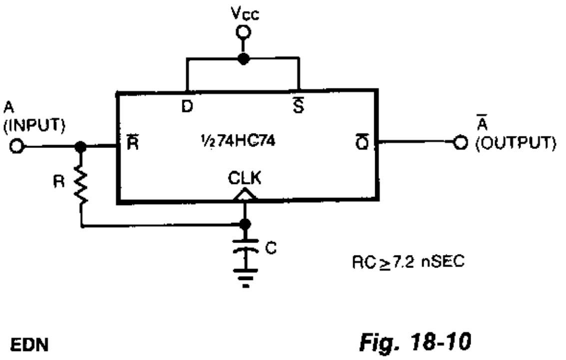
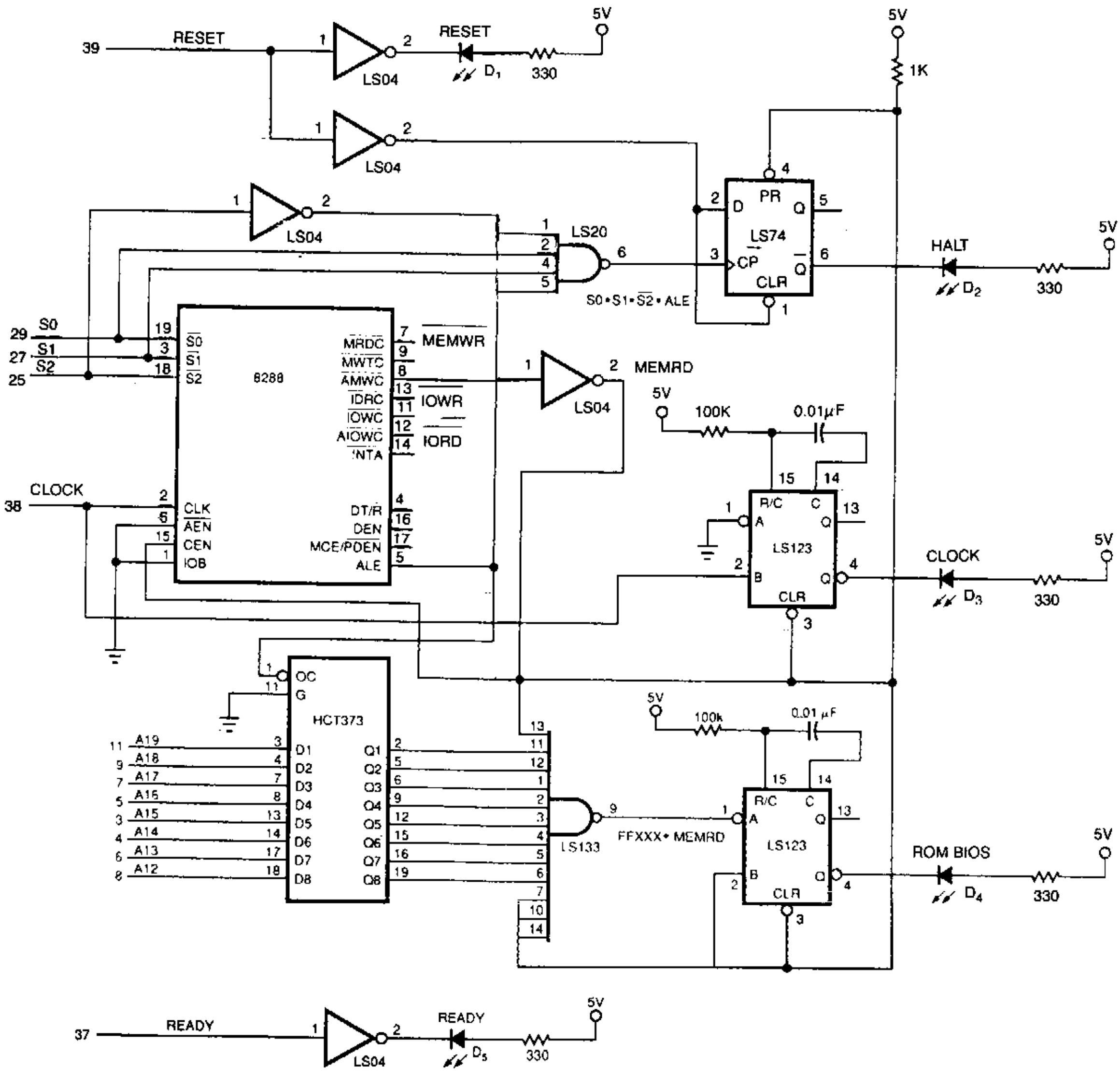


Fig. 18-10

The circuit uses one-half of a dual D flip-flop as an inverter. When the input decreases, the flip-flop resets, and its Q output increases. When the input increases, the reset line is released and Q gets clocked low. The rc delay between applying the input signal to the flip-flop's reset input and its clock input enables clocking the flip-flop on the input's positive edge. A 74HC74 dual D flip-flop, for example, requires a minimum recovery time of 5 ns after releasing the reset input before strobing its clock input. Therefore, speccing rc at greater than 7.5 ns provides adequate margin. The slight slowing of the clock edge presents no problem, because the clock input's maximum allowable rise time is a much longer 500 ns. To prevent skewing of the output's symmetry, limit the maximum input frequency to less than 10 MHz.

COPROCESSOR SOCKET DEBUGGER



EDN

Fig. 18-11

The IBM PC debugger plugs into the PC's math-coprocessor socket. The 8288 bus controller, IC1, regenerates control signals from the processor's status signals, S0, S1, and S2. Reset LED D1 lights if reset is active and holding the processor. Clock status LED D3 indicates that the processor is receiving a toggling clock signal. The address-decode logic detects when the processor is doing a jump-on-reset to the PC's BIOS ROM's power-on; self-test then detects a fatal error and halts the processor.

20-MHz-TO-NUBUS CLOCK PHASE LOCK

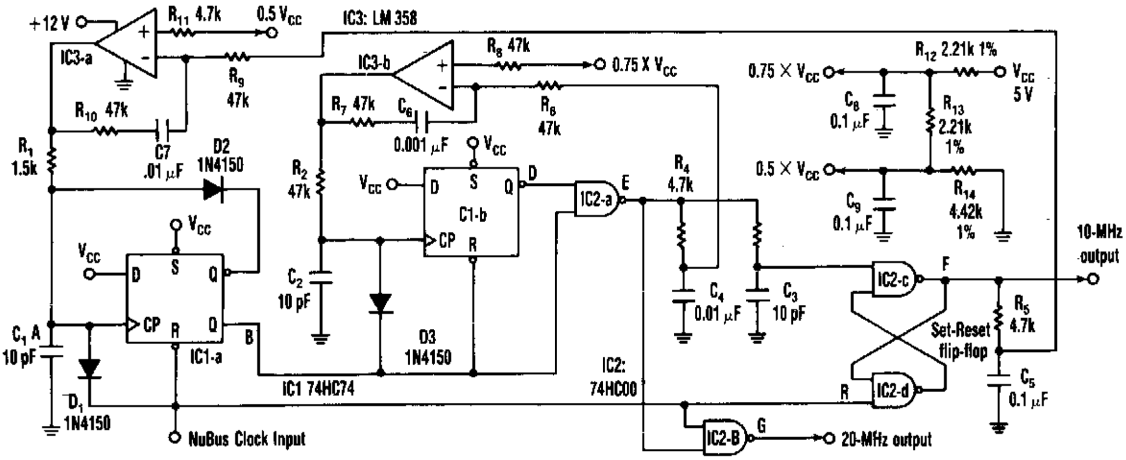
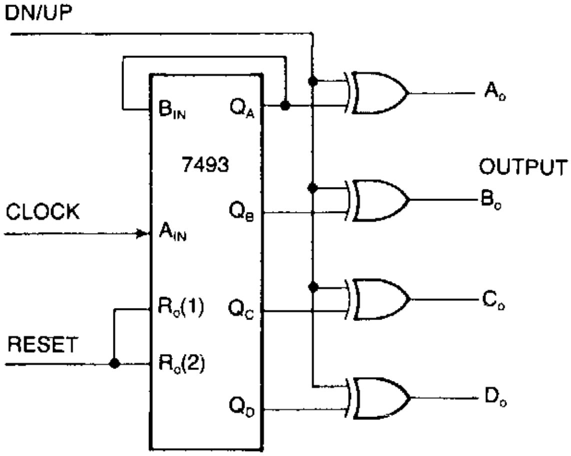


Fig. 18-12

ELECTRONIC DESIGN

The 20-MHz clock phase-locks to Apple's Mac II 10-MHz NuBus clock. It uses a simple, inexpensive CMOS circuitry to generate 10- and 20-MHz square waves. The output duty cycle settings are insensitive to V_{CC} variations. The input to the circuit is a NuBus clock signal with specifications that call for a 75 percent duty cycle at 10 MHz—a square wave that's high for 75 ns and low for the remaining 25 ns. To generate the 20-MHz signal, the circuit produces a 25-ns negative-going pulse, delayed 50 ns from the falling edge of the 10-MHz NuBus clock input at point E. NORing that pulse with the NuBus clock produces the 20-MHz clock at point G. Finally, applying the 25-ns pulse to the set input of a set-reset input, results in a 10-MHz square wave at F.

XOR GATE UP/DOWN COUNTER



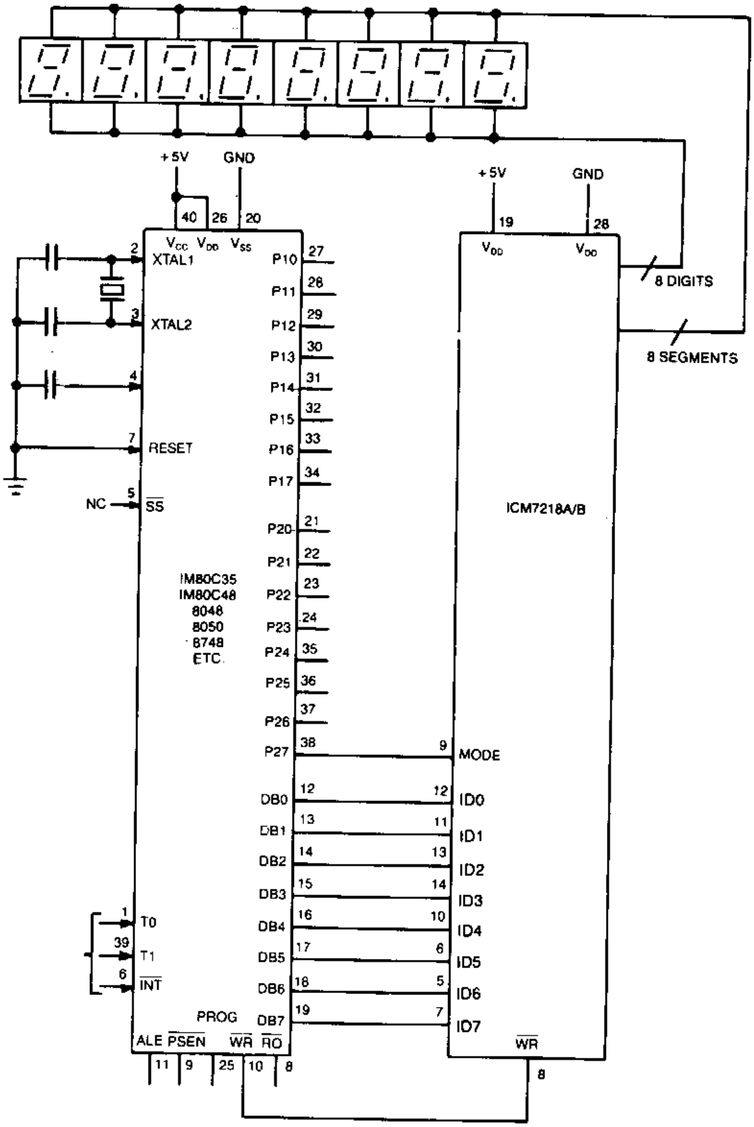
EDN

Fig. 18-13

One can transform an ordinary binary counter, such as a 7493, into an up/down counter with mode control by adding XOR gates 7486 to the counter's outputs. The circuit counts up when the DN/UP line is low and down when the DN/UP line is high.

To use the 7493 counter to count out its maximum count length of 0 – 15, connect the Q_A output to the B_{IN} input and apply clock pulses to the A_{IN} input. The reset input, when high, inhibits the count inputs and simultaneously returns the outputs A_o through D_o to low in the up-count mode or 15 in the down-count mode. For normal counting, the reset input must be low. One can easily cascade this counter by feeding the Q_D line to the clock input of a succeeding counter.

EIGHT-DIGIT MICROPROCESSOR DISPLAY



INTERSIL

Fig. 18-14

Display interface uses the ICM7218A/B with an 8048 family microcontroller. The 8-bit data bus (DB0/DB7 – ID0/ID7) transfers control and data information to the 7218 display interface on successive WRITE pulses. The mode input pins on the microcontroller. When mode is high, a control word is transferred; when mode is low, data is transferred. Sequential locations in the 8-byte static memory are automatically loaded on each successive WRITE pulse. After eight WRITE pulses have occurred, further pulses are ignored until a new control word is transferred.

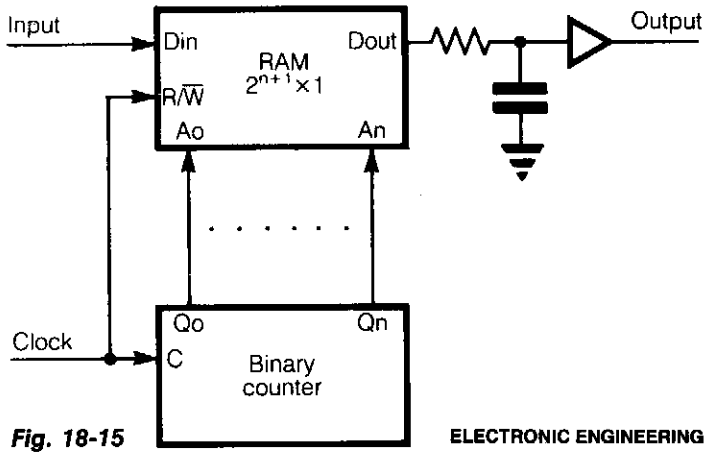


Fig. 18-15

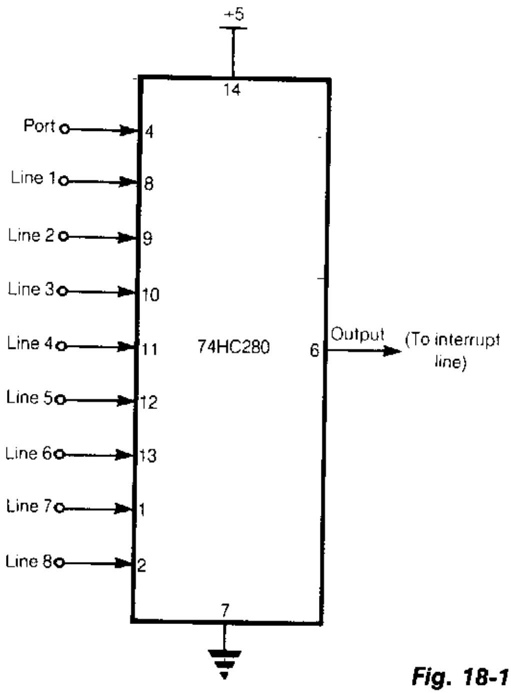
ELECTRONIC ENGINEERING

An extremely long delay of logic signal can be accomplished with this circuit. The logic signals to be delayed are applied to the D_{IN} of RAM. Address lines *Ao*, *A*1, . . . *An* are connected to outputs *Qo*, *Q*1, . . . *Qn* of a binary counter. Clock input of counter and R/$\overline{W}$ input of RAM are joined together. However, it is sometimes necessary to put an inverter between those inputs, depending on the RAM and counter employed in line. In the first half of clock interval, content on outputs of counter is increased for 1 and content of chosen memory cell is read; in the second half of the clock interval, new content from D_{IN} in the same memory cell is written. When full cycles of counting reaches the same memory cell, again we can read, in the first half of the clock interval of the following counting cycles, the chosen content. Delay time is:

$$Td = 2^{n+1} \cdot tcl$$

If clock frequency is not a multiple of input signal frequency, distortion of input signal is proportioned to the clock period. But if the clock frequency is a multiple of input signal frequency, there is no distortion. If we use RAM organized according to $2^{n+1} \times 4$ with separated data inputs and data outputs, we can have four parallel long delay lines. The resistor, capacitor, and buffer on D_{OUT} of RAM are used to save output signal in writing time, when output of RAM becomes high impedance.

LOGIC LINE MONITOR



ELECTRONIC ENGINEERING

Fig. 18-16

This circuit requires only one CMOS IC, which is available in a 14-pin surface-mount package. The figure shows the logic lines going to a standard 9-bit parity comparator chip. This device is conventionally used in data transmission and recording applications to provide a means of error-detection by comparing the received eight- or nine-bit words with their corresponding parity bits. If the sum of the one's in a received word is odd but the odd-parity bit is low, then that word is known to be in error and requires retransmission.

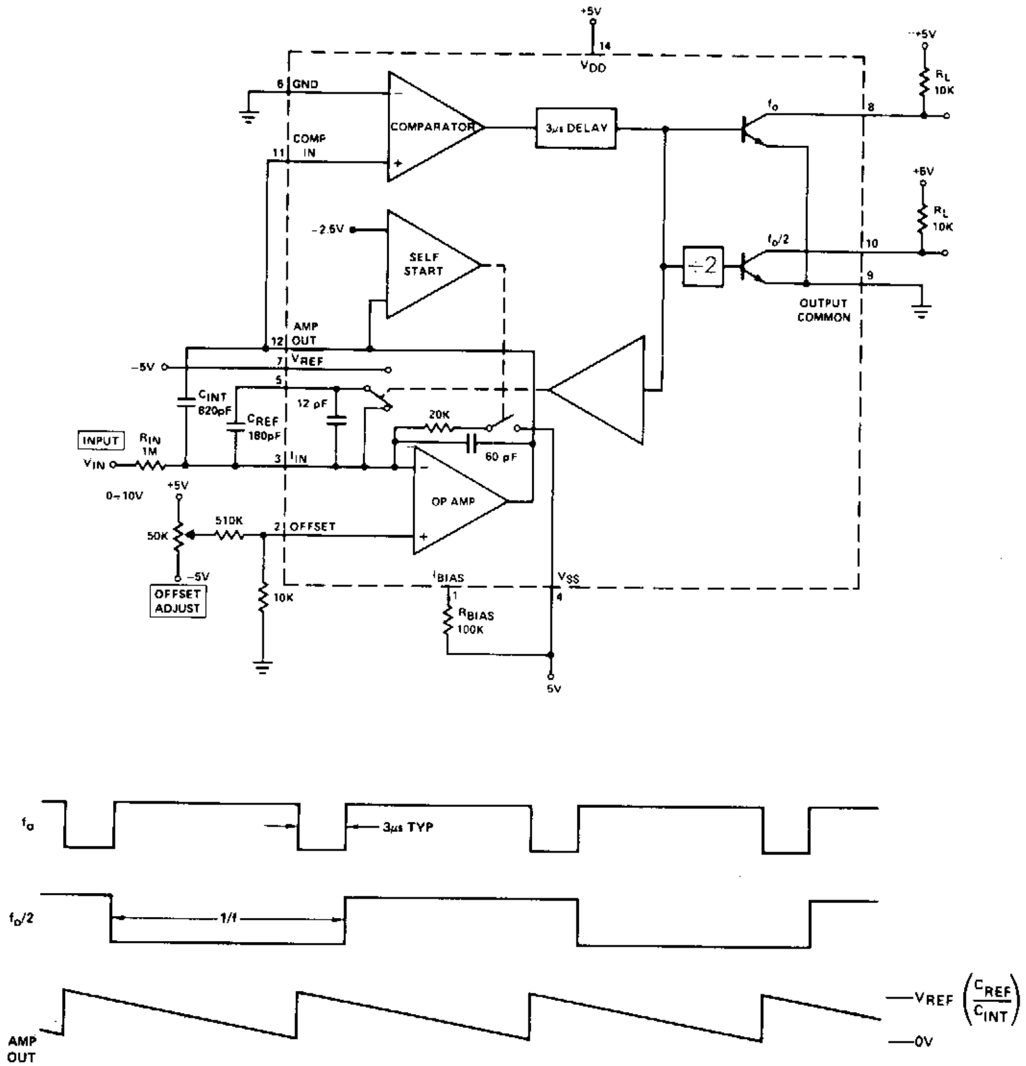
When one of the logic lines decreases, the output of the parity comparator decreases, generating a *wake-up* interrupt to the microprocessor. The ninth line comes from a port on the microprocessor and is toggled to reset the output signal high again, ready for the next logic change.

19

Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

10 Hz-TO-10kHz V/F CONVERTER



Notes:

1. To adjust f_{min}, set V_{IN} = 10 mV and adjust the 50 k offset for 10 Hz out.
2. To adjust f_{max}, set V_{IN} = 10 V and adjust R_{IN} or V_{REF} for 10 kHz out.
3. To increase f_{OUT} MAX to 100 kHz change C_{REF} to 27 pF and C_{INT} to 75 pF.
4. For high performance applications use high stability components for R_{IN}, C_{REF}, V_{REF} (metal film resistors and glass film capacitors). Also separate the output ground (Pin 9) from the input ground (Pin 6).

Fig. 19-1

LOW-FREQUENCY CONVERTER

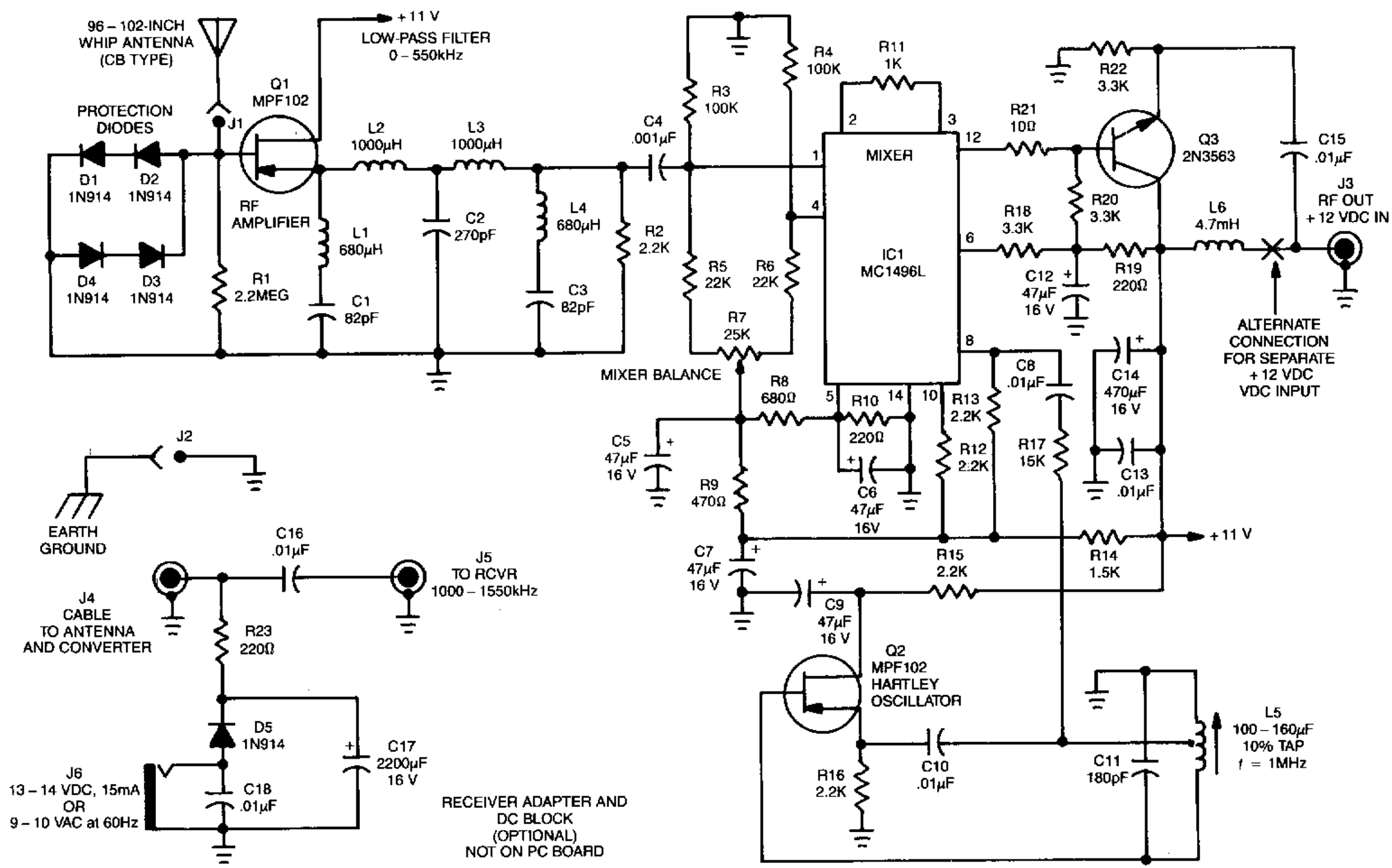


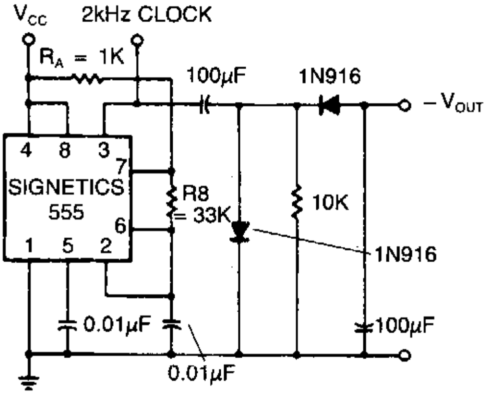
Fig. 19-2

Among the signals below 550 kHz are maritime mobile, distress, radio beacons, aircraft weather, European Longwave-AM broadcast, and point-to-point communications. The low-frequency converter converts the 10 to 500 kHz LW range to a 1010 to 1550 kHz MW range, by adding 1000 kHz to all received signals. Radio calibration is unnecessary because signals are received at the AM-radio's dial setting, plus 1 MHz; a 100-kHz signal is received at 1100 kHz, a 335-kHz signal at 1335 kHz, etc. The low-frequency signals are fed to IC1, a doubly-balanced mixer.

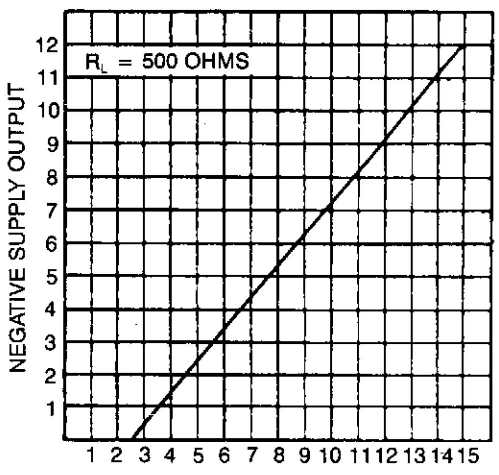
Transistor Q2 and associated circuitry form a Hartley 1000-kHz local oscillator, which is coupled from Q2's drain, through C8, to IC1 pin 8. Signals in the 10 – 550 kHz range are converted to 1010 – 1550 kHz. The mixer heterodynes the incoming low-frequency signal and local-oscillator signal. Transistor Q3 reduces IC1's high-output impedance to about 100 Ω to match most receiver inputs. Capacitor C15 couples the 1010 – 1550 kHz frequencies from Q3's emitter to output jack J3, while blocking any dc bias.

Inductor L6 couples the dc voltage that's carried in the rf signal cable from the rcvr/dc adaptor. The dc voltage and rf signals don't interfere with one another; that saves running a separate power-supply wire, which simplifies installation at a remote location. Capacitors C14 and C13 provide dc supply filtering. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

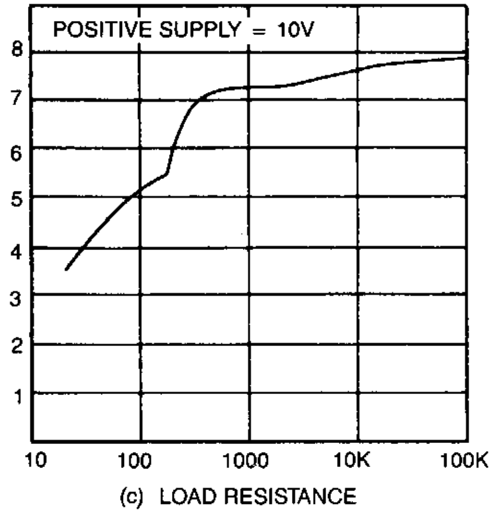
POSITIVE-TO-NEGATIVE CONVERTER



(a) POSITIVE-TO-NEGATIVE CONVERTER



(b) POSITIVE SUPPLY

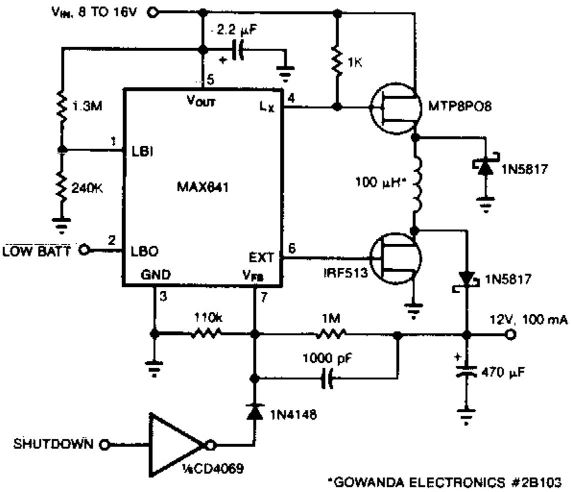


SIGNETICS

Fig. 19-3

The transformerless dc-dc converter derives a negative supply voltage from a positive. As a bonus, the circuit also generates a clock signal. The negative output voltage tracks the dc-input voltage linearity (a), but its magnitude is about 3 V lower. Application of a 500-Ω load, (b), causes 10% change from the no-load value.

BUCK/BOOST CONVERTER

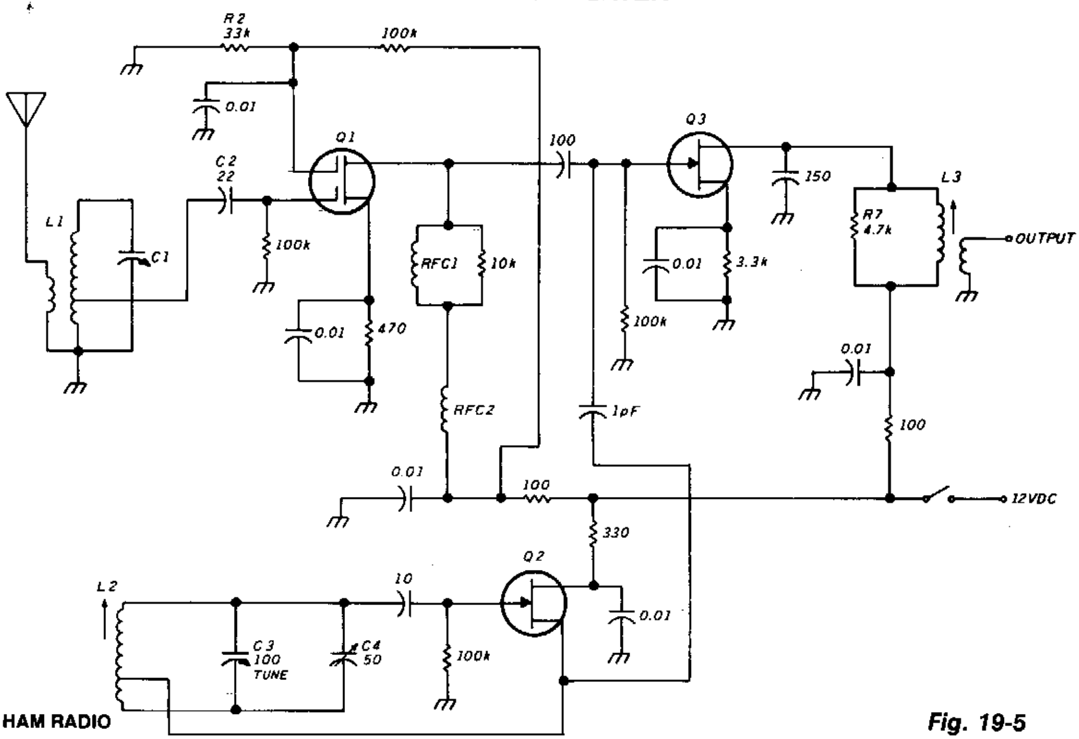


MAXIM

Fig. 19-4

This converter can accommodate wide input-voltage swings, such as the 8 to 15-V swing typical of a 12-V sealed lead/acid battery. The low battery output indicates when input voltage drops below 8 V. Pulling shutdown turns off the circuit.

4 – 18 MHz CONVERTER

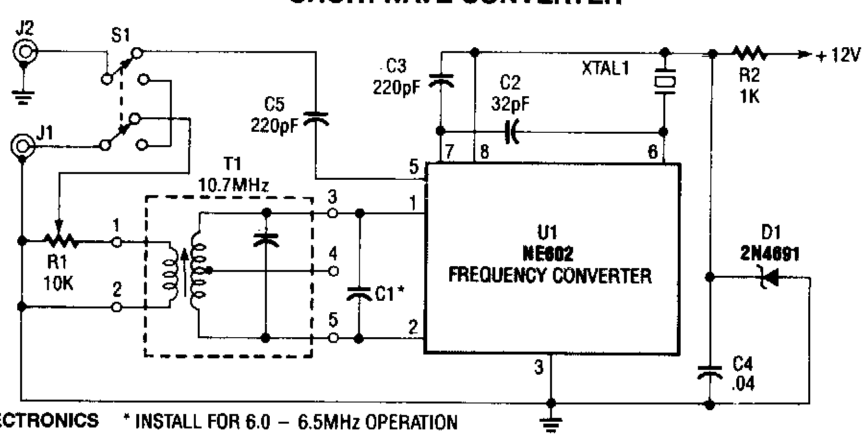


HAM RADIO

Fig. 19-5

The unit consists of rf amplifier Q1, local oscillator Q2, and mixer Q3. The two bands are covered without a bandswitch by using an i-f or 3.5 MHz. The oscillator range is 7.5 to 14.5 MHz. Incoming signals from 4 to 11 MHz are mixed with the oscillator to produce the 3.5-MHz i-f. Signals from 11 to 18 MHz are mixed with the oscillator to also produce an i-f of 3.5 MHz. At any one oscillator frequency, the two incoming signals are 7 MHz apart. Rf amplifier input C1/L1 comprises a high-*Q*, lightly loaded, tuned circuit; this is essential for good band separation.

SHORTWAVE CONVERTER



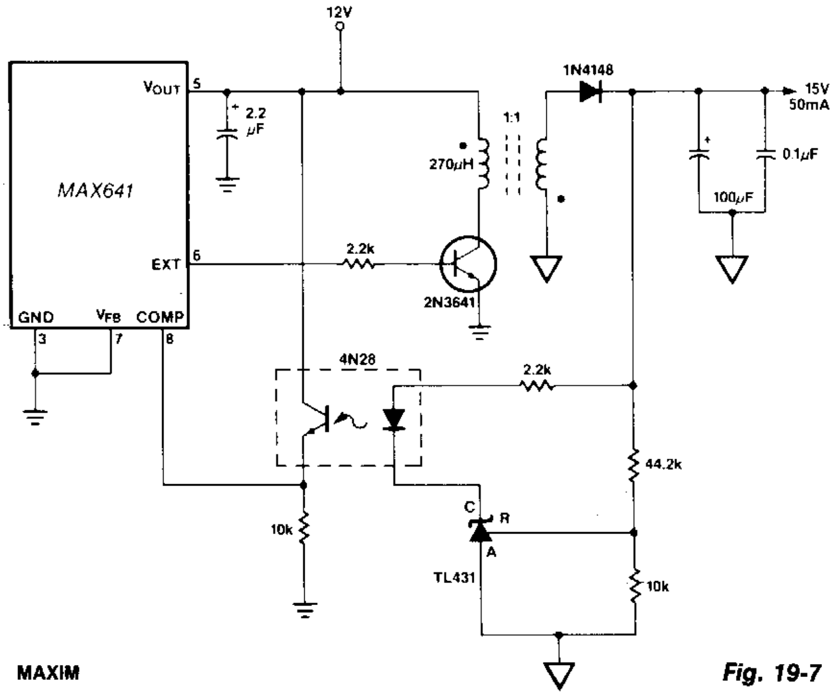
POPULAR ELECTRONICS * INSTALL FOR 6.0 – 6.5MHz OPERATION

Fig. 19-6

SHORTWAVE CONVERTER (*Cont.*)

The NE602, U1, contains oscillator and mixer stages. The mixer combines the oscillator signal with the input rf signal to produce signals whose frequencies are the sum and difference of the input frequencies. For example, a 7.5-MHz signal is picked up by the antenna and mixes with the 8.5-MHz oscillator frequency. The difference between those two signals is 1 MHz—right in the center of your AM dial. Transformer T1 is a 10.7-MHz i-f transformer.

ISOLATED +15 V DC-DC CONVERTER

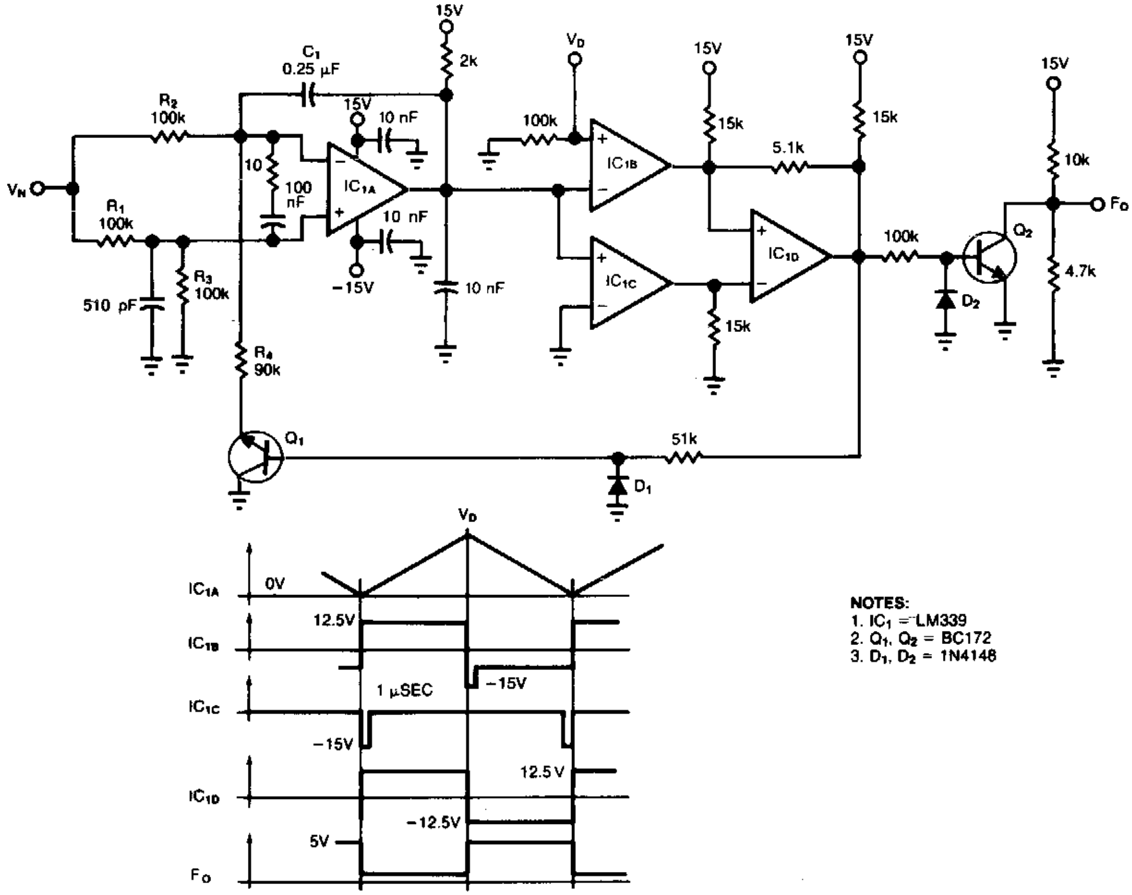


MAXIM

Fig. 19-7

In this circuit, a TL431 shunt regulator is used to sense the output voltage. The TL431 drives the LED of a 4N28 optocoupler which provides feedback to the MAX641 while maintaining isolation between the input, +12 V, and the output, +15 V. In this circuit, the +15 V output is fully regulated with respect to both line and load changes.

VOLTAGE RATIO-TO-FREQUENCY CONVERTER

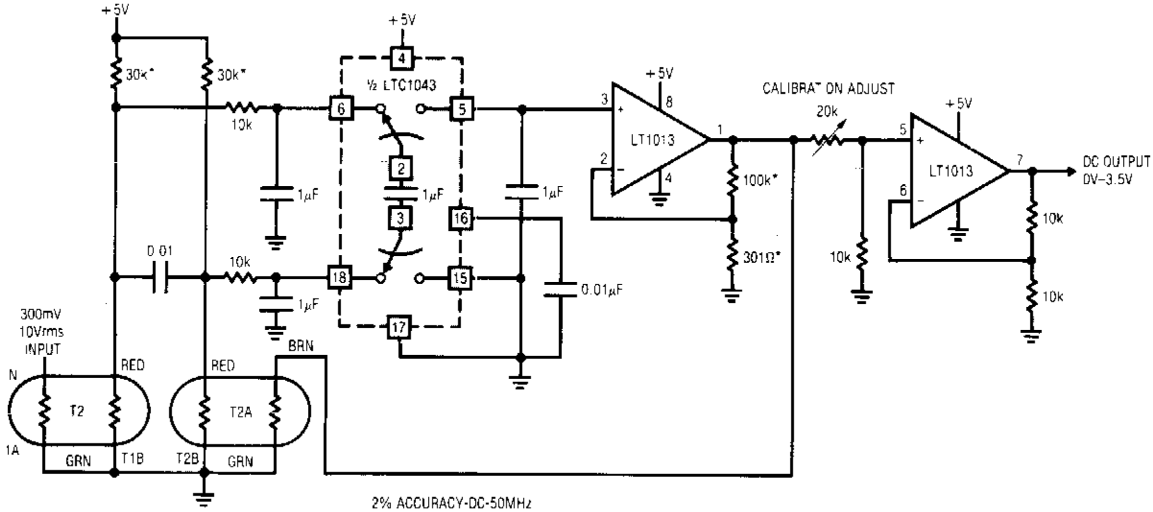


EDN

Fig. 19-8

The circuit accepts two positive-voltage inputs V_N and V_D and provides a TTL-compatible output pulse train whose repetition rate is proportional to the ratio V_N/V_D. Full-scale output frequency is about 100 Hz, and linearity error is below 0.5 percent. The output F_o equals KV_N/V_D, where $K = 1/(4R_2C_1)$ provided $R_1 = R_3$. Op amp IC1A alternately integrates $V_N/2$ and $-V_N/2$, producing a sawtooth output that ramps between the V_D level and ground. When transistor Q1 is on, for example, IC1A integrates $-V_N/2$ until its output equals V_D. At that time, the IC1B comparator switches low, causing IC1D's bistable output to go low, which turns off Q1. IC1A's output then ramps in the negative direction. When the output reaches 0 V, the IC1C comparator switches, Q1 turns on, and the cycle repeats. Transistor Q2 converts the IC1D output to TTL-compatible output logic levels. Setting V_D to 1.00 V yields a linear voltage-to-frequency converter ($F_o = KV_N$), and setting V_N to 1.00 V yields a reciprocal voltage-to-frequency converter ($F_o = KV_D$).

50-MHz THERMAL RMS-TO-DC CONVERTER



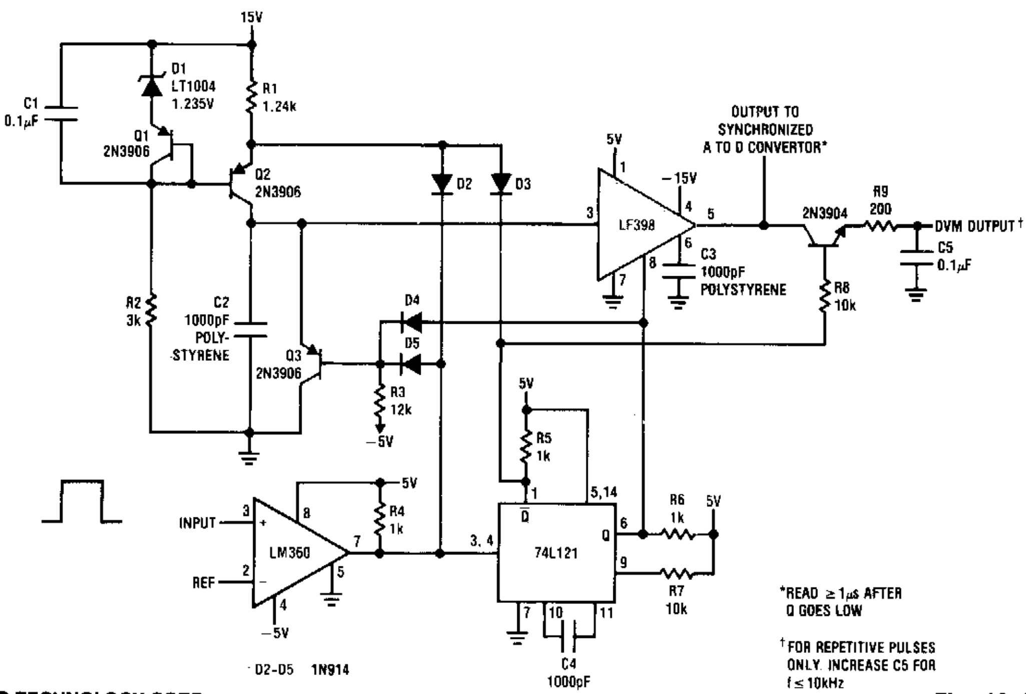
2% ACCURACY-DC-50MHz
100:1 CREST FACTOR CAPABILITY
T1–T2 = YELLOW SPRINGS INST. CO. THERMISTOR COMPOSITE
ENCLOSE T1 AND T2 IN STYROFOAM

*1% RESISTOR

LINEAR TECHNOLOGY CORP.

Fig. 19-9

PULSE WIDTH-TO-VOLTAGE CONVERTER



D2–D5 1N914

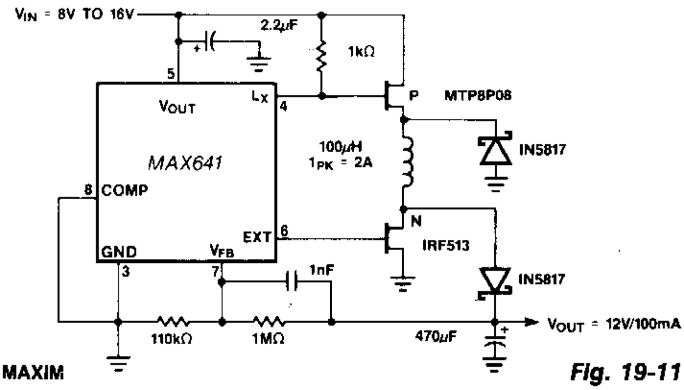
*READ ≥ 1μs AFTER Q GOES LOW

†FOR REPETITIVE PULSES ONLY. INCREASE C5 FOR f ≤ 10kHz

LINEAR TECHNOLOGY CORP.

Fig. 19-10

STEP UP/DOWN DC-DC CONVERTER



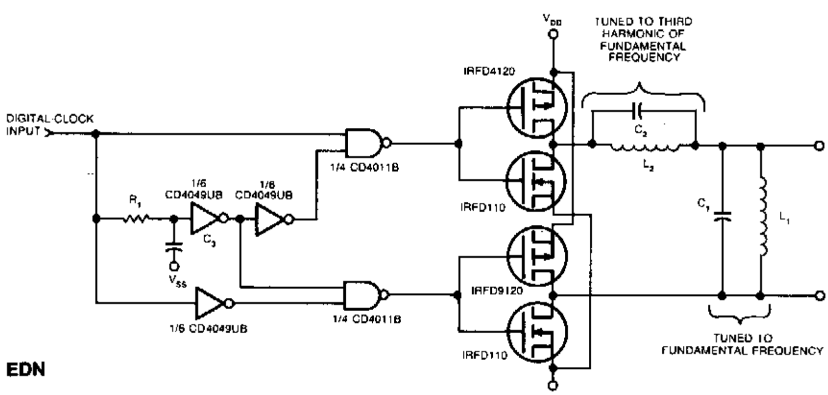
MAXIM

Fig. 19-11

Positive output step-up and step-down dc-dc converters have a common limitation in that neither can handle input voltages that are both greater than or less than the output. For example, when converting a 12-V sealed lead/acid battery to a regulated +12 V output, the battery voltage might vary from a high of 15 V down to 10 V.

By using a MAX641 to drive separate P- and N-channel MOSFETs, both ends of the inductor are switched to allow noninverting buck/boost operation. A second advantage of the circuit over most boost-only designs is that the output goes to 0 V when shutdown is activated. Inefficiency is a drawback because two MOSFETs and two diodes increase the losses in the charge and discharge path of the inductor. The circuit delivers +12 V at 100 mA at 70 percent efficiency with an 8-V input.

SQUARE-TO-SINE WAVE CONVERTER

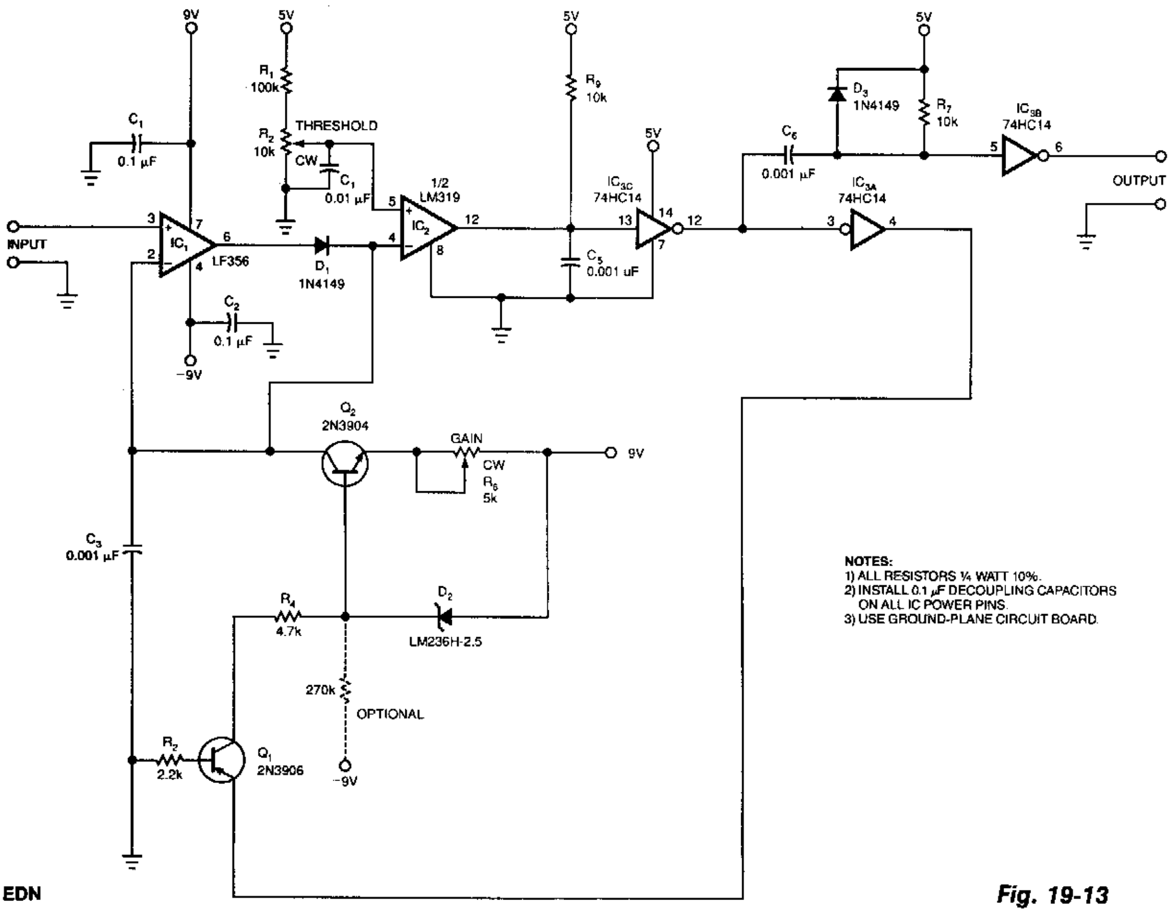


EDN

Fig. 19-12

Two pairs of MOSFETs form a bridge that alternately switches current in opposite directions. Two parallel-resonant LC circuits complete the converter. The L1/C1 combination is resonant at the fundamental frequency; the L2/C2 combination is resonant at the clock frequency's third harmonic and acts as a trap. T1 and C3 ensure that both halves of the MOSFET bridge are never on at the same time by providing a common delay to the gate drive of each half. Select the values of R1 and C3 to yield a time constant that's less than 5% of the clock's period. You can add an output amplifier for additional buffering and conditioning of the circuit's sine-wave output.

PULSE HEIGHT-TO-WIDTH CONVERTER



EDN

Fig. 19-13

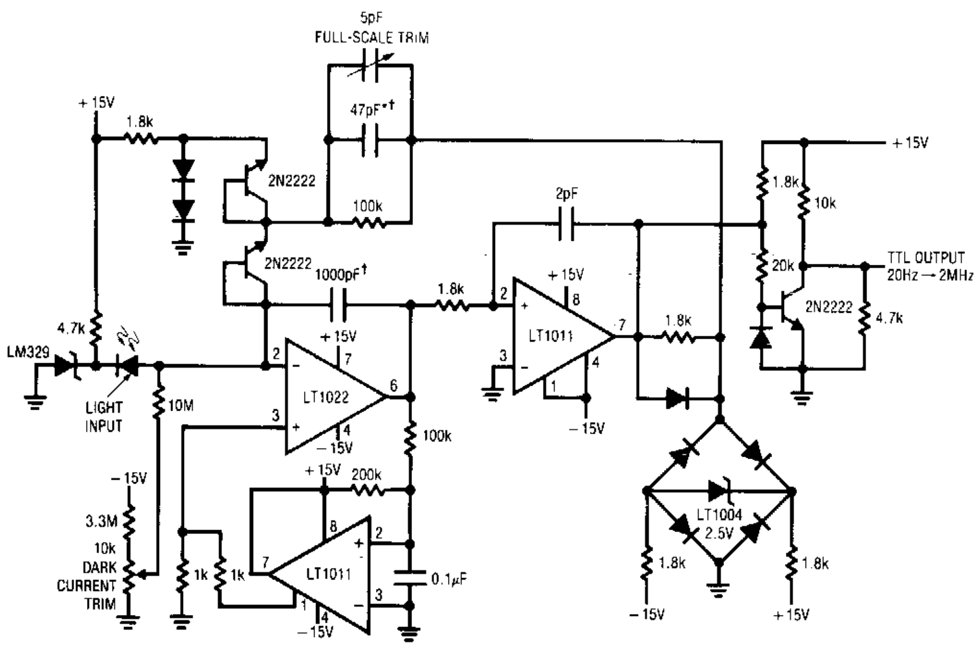
The output-pulse width from the circuit is a linear function of the input pulse's height. You can set the circuit's input threshold to discriminate against low-level pulses, while fixed components limit the circuit's maximum output-pulse width.

With a 270-KΩ resistor connected from the −9 V supply to the base lead of Q2, this circuit can handle input pulses separated by 20 μs for correct operation. The turn-off time of zener diode D2 sets the lower limit for the input-pulse repetition rate.

IC1, D1, and C3 detect the peak of the input pulse. The comparator IC2, triggers at your preset threshold. The RC delay network, R9 and C5, hold off inverter IC3's changing state until the completion of peak detection. After IC3A changes state, Q1 turns on and then turns on Q2, a constant-current source.

Constant-current source Q1 then discharges C3, the peak-detecting capacitor. When C3 has discharged below IC2's threshold, IC2's output decreases, as do pins 6 and 4 of IC3. The output-pulse width is a function of this discharge time, which you can adjust with R6. C6 and R7 control the maximum output-pulse width, which is 8 μs max.

PIN PHOTODIODE-TO-FREQUENCY CONVERTER



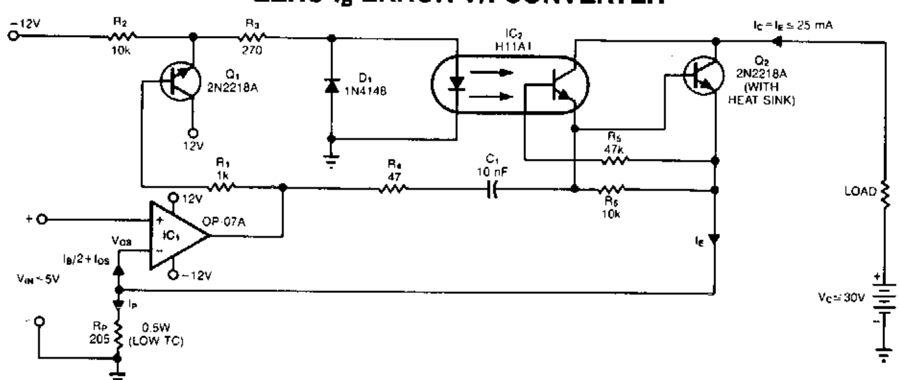
SCALE FACTOR =
1nW/Hz AT 900 NANOMETERS FROM 20nW TO 2mW
= HEWLETT PACKARD PHOTODIODE HP5082-4204

= 1N4148
†POLYSTYRENE
*SELECT VALUE FOR 2mW IN = 2MHz OUT.

LINEAR TECHNOLOGY CORP.

Fig. 19-14

ZERO I_B ERROR V/I CONVERTER

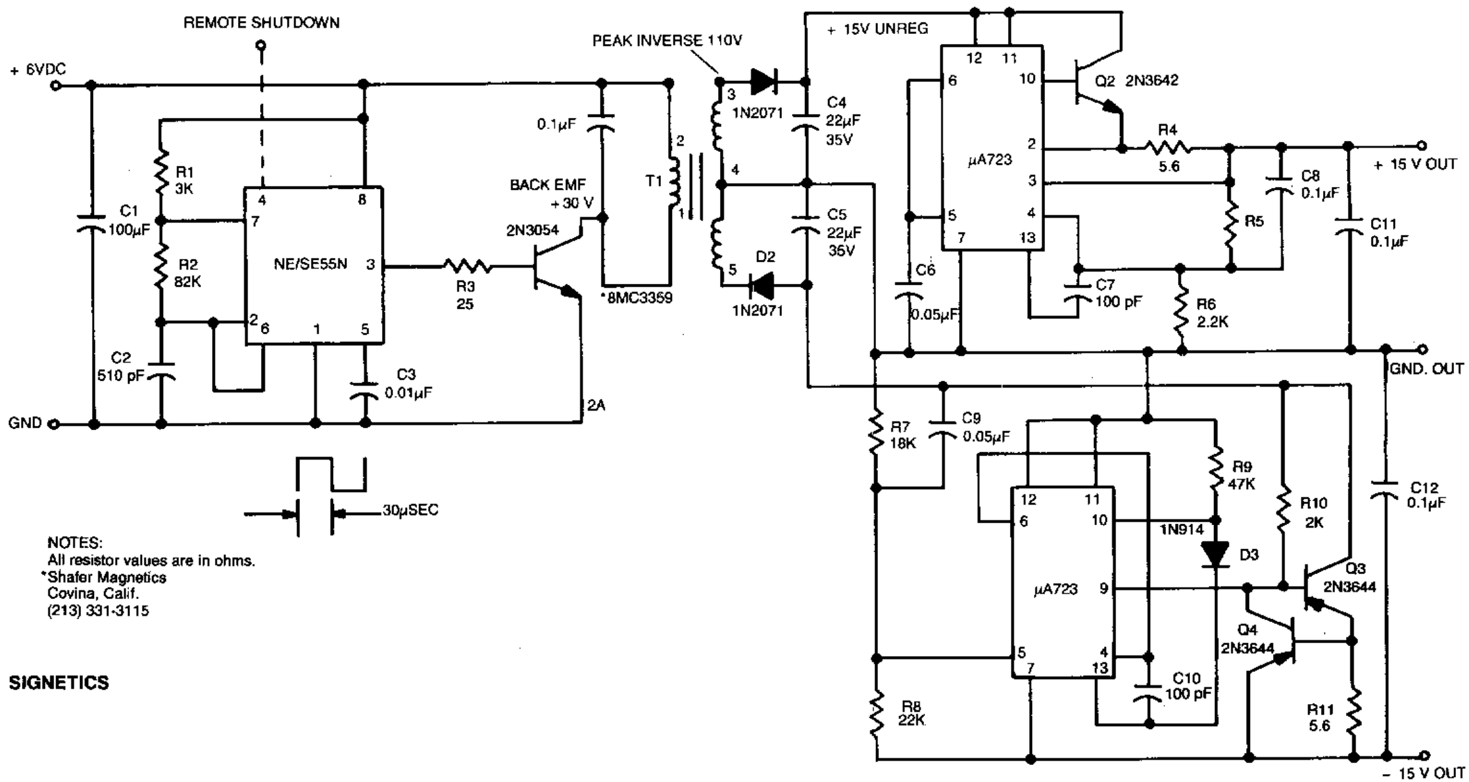


EDN

Fig. 19-15

Single programming resistor R_p provides an output-current range of about six decades. Note that this resistor's TC is also a potential source of error; it dissipates 125 mW when V_{IN} = 5 V. The maximum deviation is typically 50 nA or 0.0002% of full scale. This voltage-controlled current source uses optocoupler IC2 to eliminate an error found in more conventional circuits and which is caused by the output transistor's base current.

REGULATED DC-DC CONVERTER



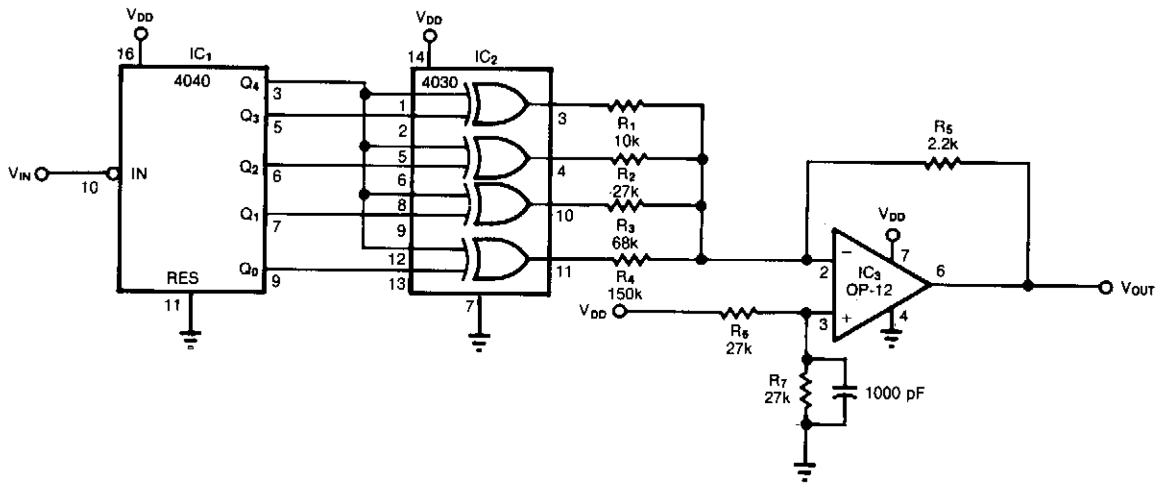
NOTES:
All resistor values are in ohms.
*Shafer Magnetics
Covina, Calif.
(213) 331-3115

SIGNETICS

Fig. 19-16

The regulated dc-dc converter produces 15-Vdc outputs from a +5 Vdc input. Line and load regulation is 0.1%.

PULSE TRAIN-TO-SINUSOID CONVERTER



EDN

Fig. 19-17

The circuit lets you convert a serial pulse stream or sinusoidal input to a sinusoidal output at 1/32 the frequency. By varying the frequency of V_{IN}, you can achieve an output range of 10^7:1—from about 100 kHz to less than 0.01 Hz. The output resembles that of a 5-bit d/a converter operating on parallel digital data.

Counter IC1 generates binary codes that repeatedly scan the range from 00000 to 11111. The output amplifier adds the corresponding XOR gate outputs, V_{DD} or ground, weighted by the values of input resistors R1 through R4. The 16 counter codes 00000 to 01111, for instance, pass unchanged to the XOR gate outputs, and cause V_{OUT} to step through the half-sinusoidal cycle for maximum amplitude to minimum amplitude.

Counter output Q4 becomes high for the next 16 codes, causing the XOR gates to invert the Q0 through Q3 outputs. As a result, V_{OUT} steps through the remaining half cycle from minimum to maximum amplitude. The counter then rolls over and initiates the next cycle. You can change the R1 through R4 values to obtain other V_{OUT} waveforms. V_{DD} should be at least 12 V to assure maximum-frequency operation from IC1 to IC2.

20

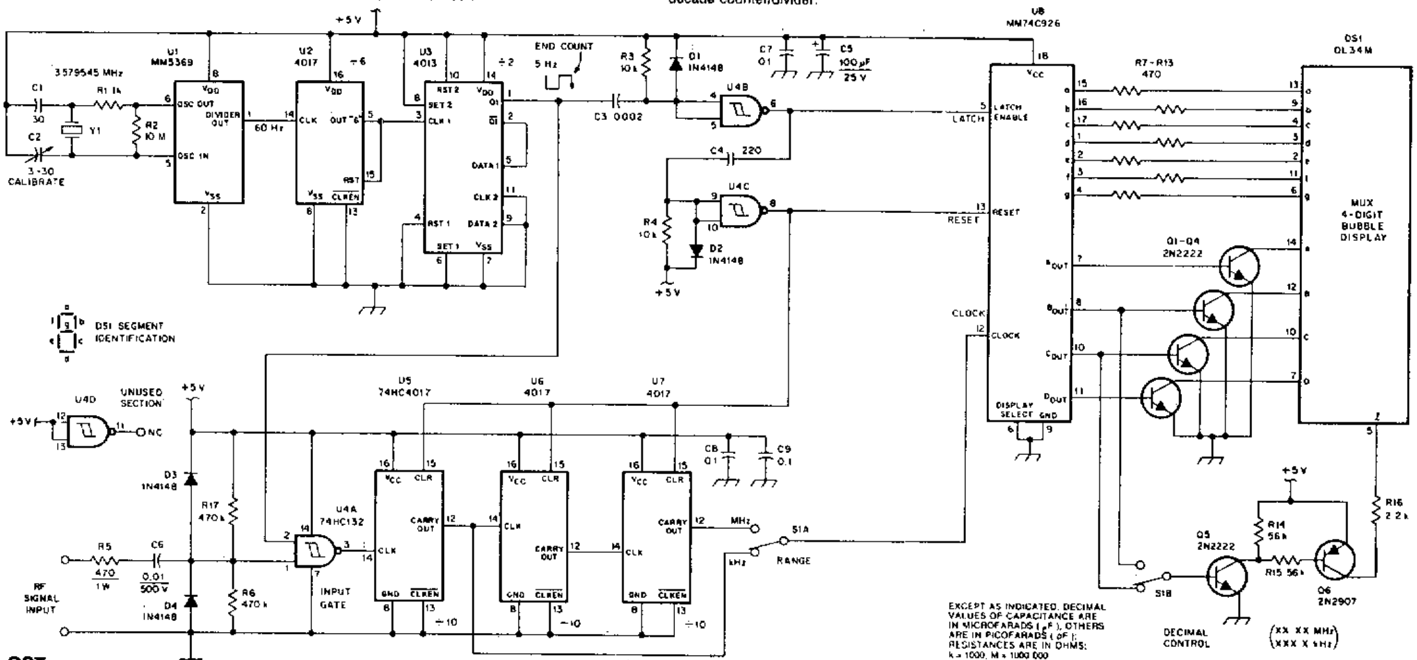
Counters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Low-Cost Frequency Counter
Up/Down Counter/Extreme Count Freezer
10-MHz Frequency Counter
Low-Power Wide-Range Programmable Counter
40-MHz Universal Counter
Frequency Counter Preamp
1.2-GHz Frequency Counter

LOW-COST FREQUENCY COUNTER

C1—30 pF NPO.
C2—3-30 pF trimmer.
D1-D4—1N4148 or equiv.
DS1—DL34M 4-digit, 7-segment, common-cathode MUX display.
Q1-Q5—2N2222, 2N3904 or equiv.
Q6—2N2907, 2N3906 or equiv.
U1—MM5369AA oscillator/divider.
U2, U6, U7—CD4017 CMOS decade counter/divider.
U3—CD4013 CMOS dual flip-flop.
U4—74HC132 HCMOS high-speed quad 2-input NAND.
U5—74HC4017 HCMOS high-speed decade counter/divider.
U8—74C926 CMOS 4-digit counter/MUX driver.
Y1—3.579545-MHz crystal.



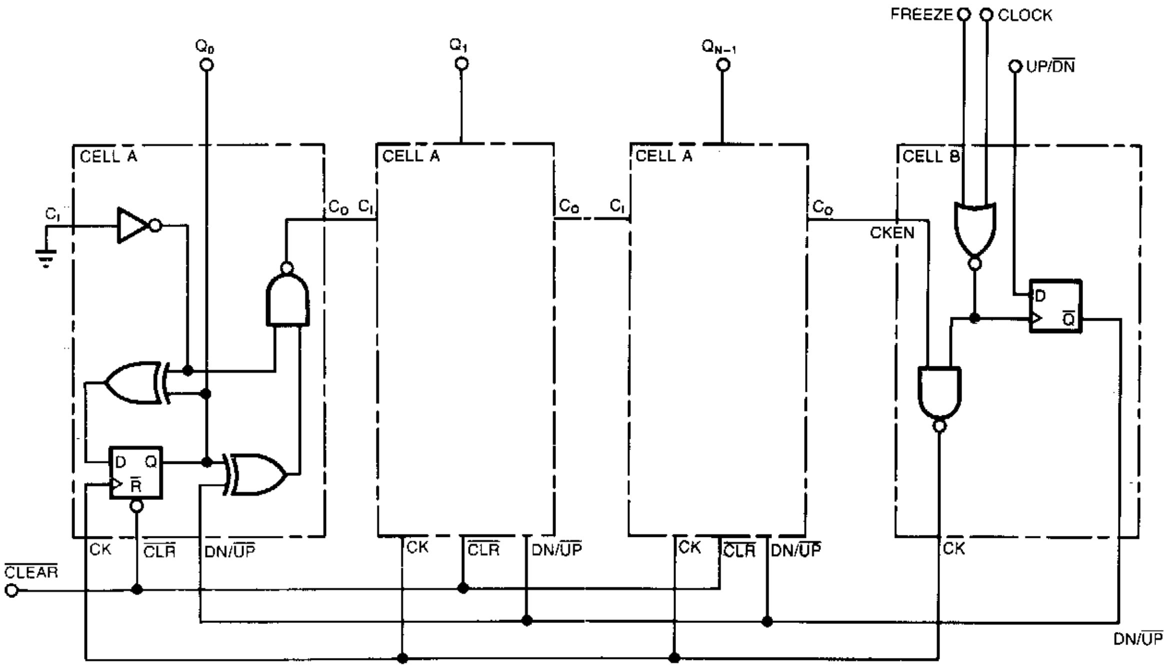
QST

Fig. 20-1

This counter uses a four-digit display, but with a flip of the range switch, it can display frequencies from 1 to 40 MHz, with a resolution of 100 Hz. The MM74C926 CMOS IC contains a four-digit decimal counter that can latch a given count and then use this information to drive a 7-segment, common-cathode multiplexed (MUX) display. The block diagram and schematic show the operation of the counter. Crystal-controlled timer U1 through U3 produces a 5-Hz square wave used for timing the frequency count. Y1 is a TV color-burst crystal operating in a reliable circuit that controls the oscillator frequency. U1 acts as the oscillator and also divides the fundamental operating frequency of 3.579545 MHz to produce a square-wave output of 60 Hz. U2 divides the 60-Hz output of U1 by six. In turn, the 10-Hz output of U2 is divided by two in U3, a dual flip-flop, to produce the 5-Hz pulse.

A quad, two-input, Schmitt-triggered NAND U4 is used for gating the rf-signal input and for generating the counter control pulses—Section U4D is unused. The 5-Hz output pulse of U3 is applied to the rf-signal input gate at U4A pin 2. When the timer output signal is low, the gate is closed.

UP/DOWN COUNTER/EXTREME COUNT FREEZER



STATE TABLE

FREEZE	UP/$\overline{DN}$	CLOCK	CURRENT STATE $Q_{N-1} \ldots Q_0$	NEXT STATE $Q_{N-1} \ldots Q_0$
L	H	↑	11. . .10	11. . .11
L	H	↑	11. . .11	11. . .11
L	L	↑	00. . .01	00. . .00
L	L	↑	00. . .00	00. . .00
H	X	X	Q_X	Q_X

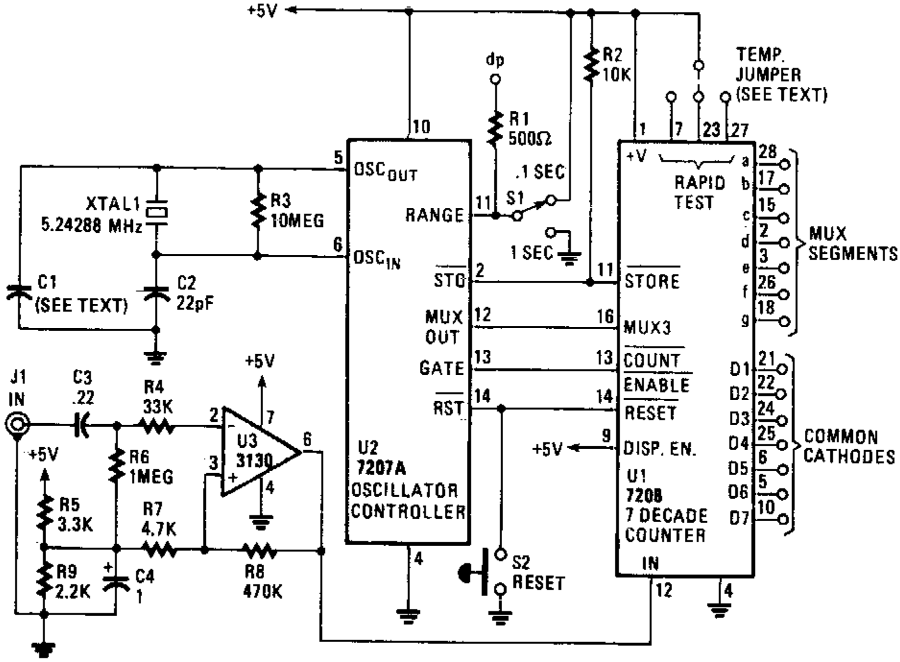
EDN

Fig. 20-2

The discrete-gate up/down-counter design has the unusual property of freezing, or saturating, when it reaches its lowest count in the down-count mode or its highest count in the count-up mode instead of rolling over and resetting as do most counters. This property proves especially useful in position-control systems. For example, you wouldn't want a robot's arm to slowly move to full extension as the counter counts up and then have it suddenly slam back to its rest position when the counter resets to zero.

You can cascade as many of the A cells as you need because the counter's outputs are synchronous. The B cell accepts the carry bit from the most significant bit's A cell and provides the clock control that stops the counter. Make sure that the freeze input to the B cell doesn't get asserted when the clock input is low; otherwise, the counter might make an extra count.

10-MHz FREQUENCY COUNTER

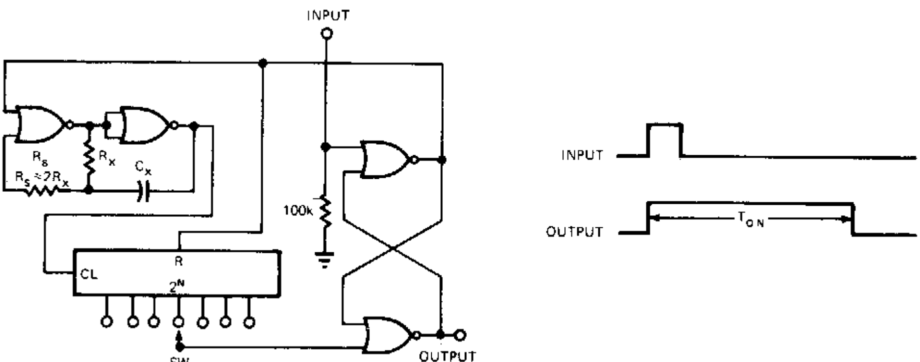


HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 20-3

The circuit consists of ICM7208 seven-decade counter U1, ICM7207A oscillator controller U2, and CA3130 biFET op amp U3. IC U1 counts input signals, decodes them to 7-segment format, and outputs signals that are used to drive a 7-digit display. IC U2 provides the timing for U1, while U3 conditions the input to U1. The 5.24288-MHz crystal frequency is divided by U2 to produce a 1280-MHz multiplexing signal at pin 12 of U2. That signal is input to U1 at pin 16 and used to scan the display digits in sequence.

LOW-POWER WIDE-RANGE PROGRAMMABLE COUNTER



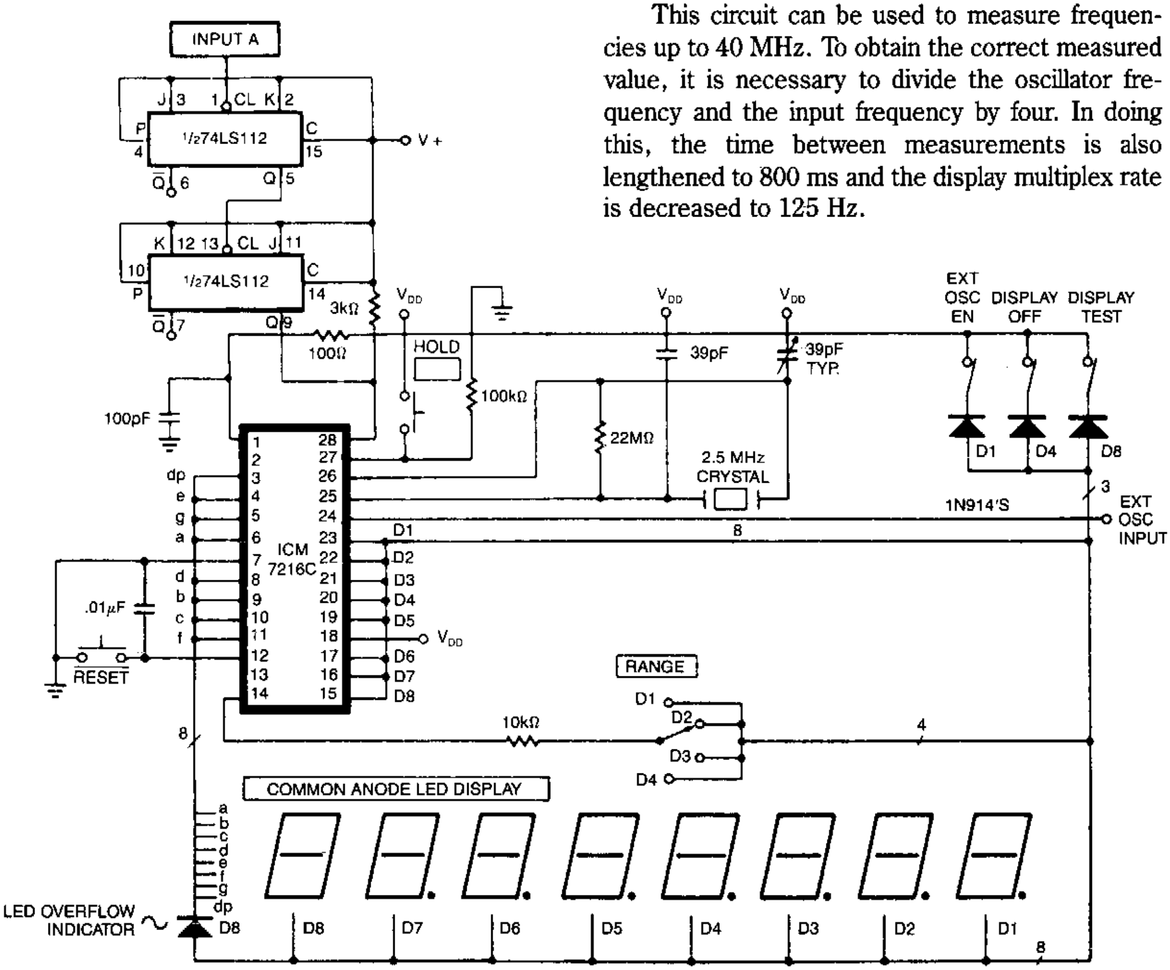
EDN

Fig. 20-4

This CMOS circuit can be used as a 1-shot time delay switch and general-purpose timer. The circuit consists of a gated oscillator and a latch made from one CD4001 quad 2-input NOR gate as shown and a CD4020 14-stage counter. T_{ON} is a function of the oscillator frequency from the $R_X C_X$ and the proper 2^N output from the counter. A pulse applied to the latch will "enable" the oscillator and counter. The latch output will remain high until the 2^N count resets the latch and disables the oscillator and counter. The circuit provides μs to hour interval timing. The extraordinarily long periods available from the CMOS oscillator, combined with the 14-stage counter, make this range possible. Further decoding is required for variations finer than a power of two.

40-MHz UNIVERSAL COUNTER

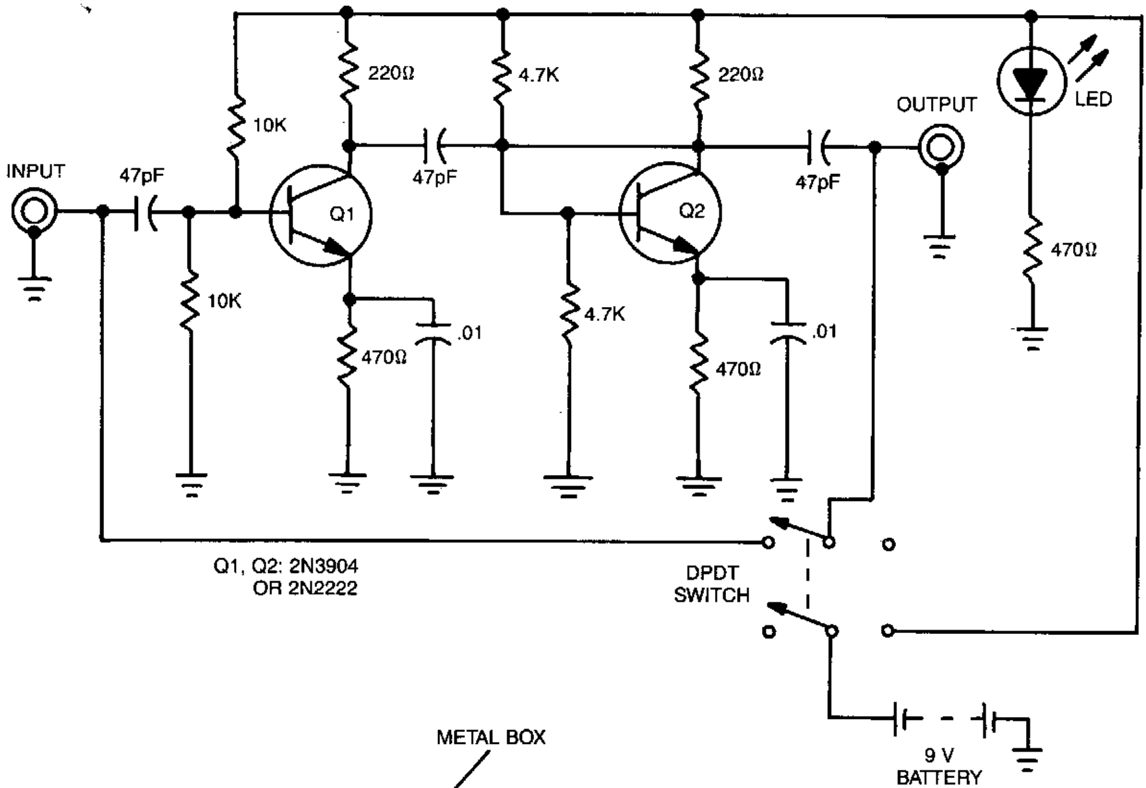
This circuit can be used to measure frequencies up to 40 MHz. To obtain the correct measured value, it is necessary to divide the oscillator frequency and the input frequency by four. In doing this, the time between measurements is also lengthened to 800 ms and the display multiplex rate is decreased to 125 Hz.

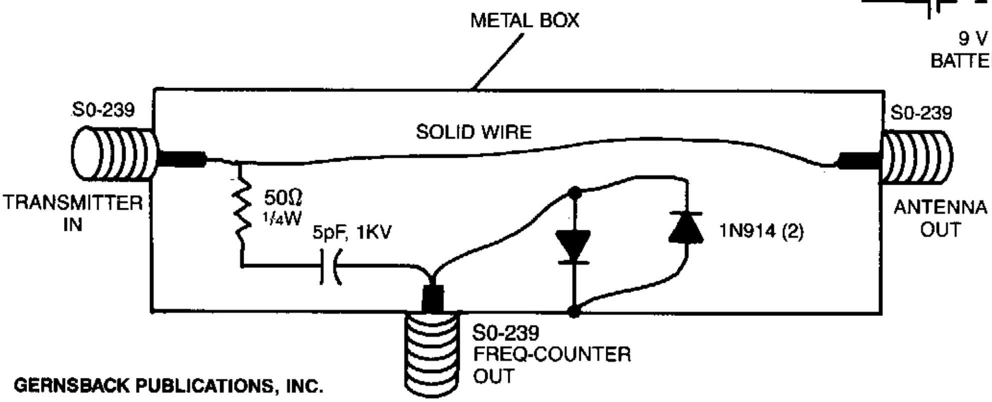


INTERSIL

Fig. 20-5

FREQUENCY-COUNTER PREAMP





GERNSBACK PUBLICATIONS, INC.

Fig. 20-6

By using the preamplifier with a short length of shielded cable and clip leads, signals that generally could not generate a readout, generate precise and stable readouts on the counter. The DPDT switch is used to bypass the circuit when amplification is not needed. The preamplifier can also be used for other purposes. For example, the unit was also tested as a receiver preamplifier and increased received signal strength about 6 *S*-units at 30 MHz. A line tap can be used to measure the frequency directly at the output of a transmitter. The entire circuit for that consists of two diodes, one resistor, and one capacitor. The line tap simply picks a low-amplitude signal for measurement by the frequency counter. The tap can be connected to transmitters with an output power of between 1 and 250 W.

1.2-GHz FREQUENCY COUNTER

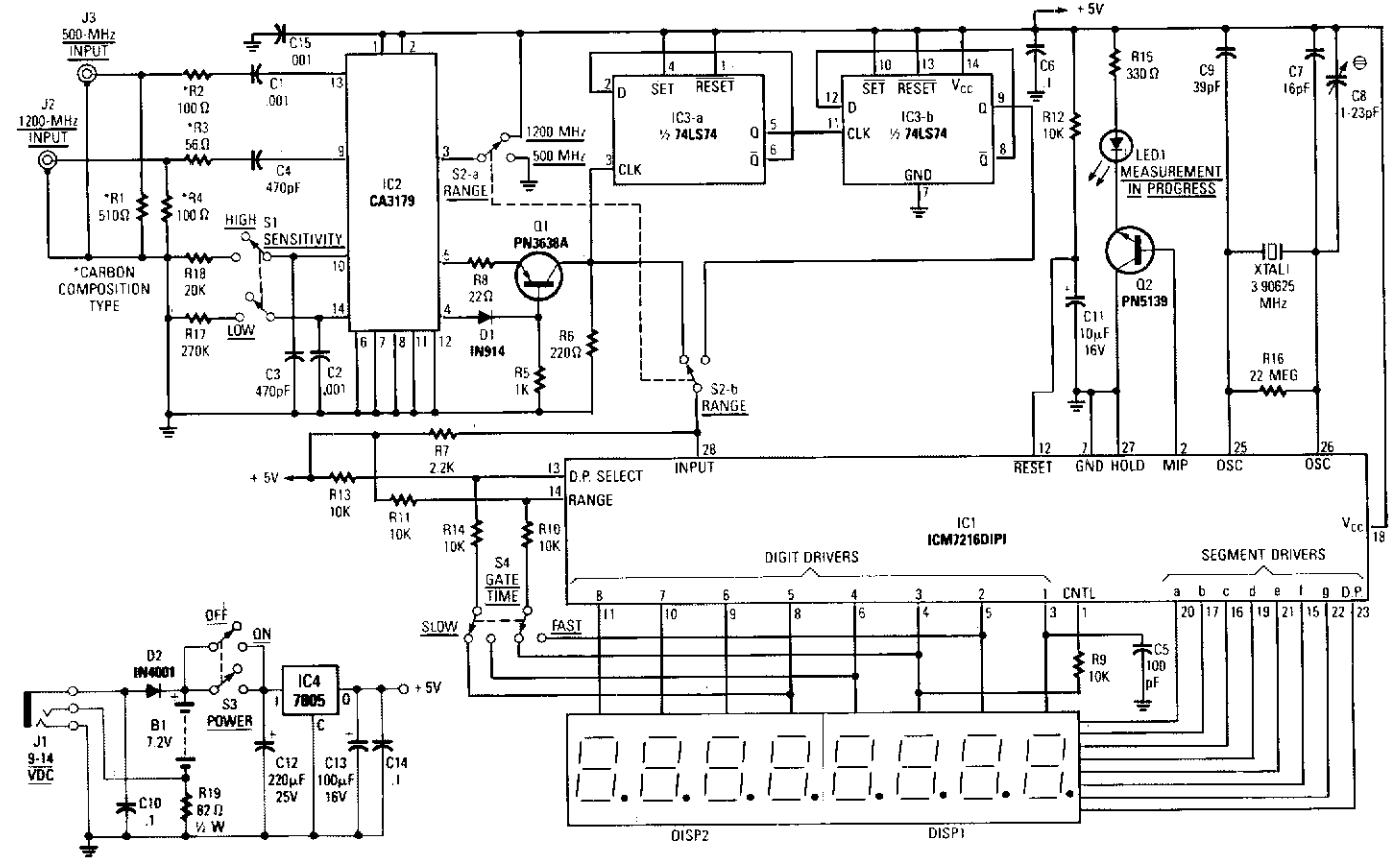


Fig. 20-7

The output of the CA3179 is fed through the D1/Q1 circuit. Those components serve to boost the 1-V output of the CA3179 to a standard TTL level. Then, depending on the position of range switch S2b, the signal is passed directly to the 7216, or through the divide-by-four circuit built from the two D flip-flops in IC3.

The other half of the range switch S2a controls the voltage at pin 3 of the CA3179. When pin 3 is high, the signal applied to pin 9 is fed through an extra internal divide-by-four stage before it is amplified and output on pins 4 and 5. When pin 3 is low, the signal on pin 13 is simply processed for output without being divided internally.

A 3.90625-MHz crystal provides the time base; the crystal yields a fast gate time of 0.256 second. The displayed frequency equals the input frequency divided by 1000 in the fast mode. In slow mode, gate time is 2.56 seconds. The displayed frequency equals the input frequency divided by 100 in the slow mode.

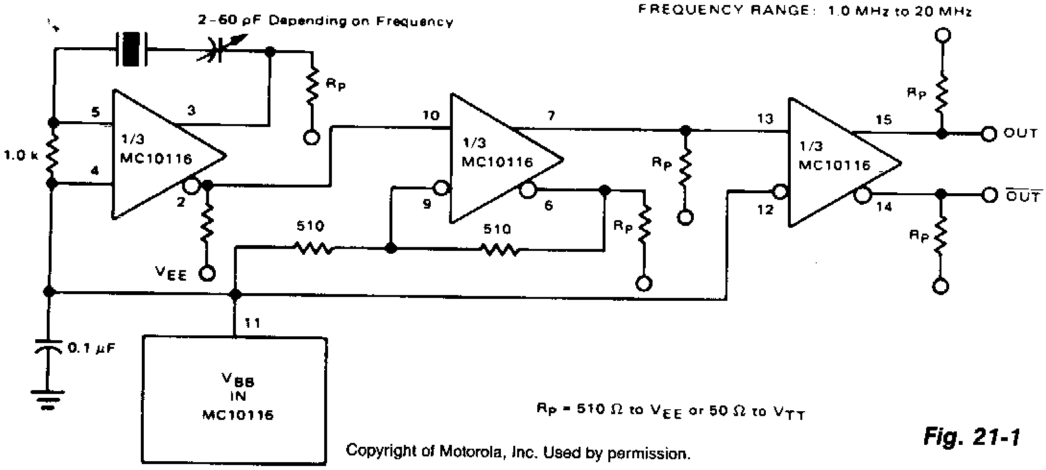
Switch S4, gate time, performs two functions. First it selects the appropriate gate time according to which digit output of IC1 the range input is connected to. Another of the 7216's inputs is also controlled by S4: the dp select input. The decimal point of the digit output to which that pin is connected will be the one that lights up. The correct decimal point illuminates, according to the position of S4, to provide a reading in MHz.

21

Crystal Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

FUNDAMENTAL-FREQUENCY CRYSTAL OSCILLATOR

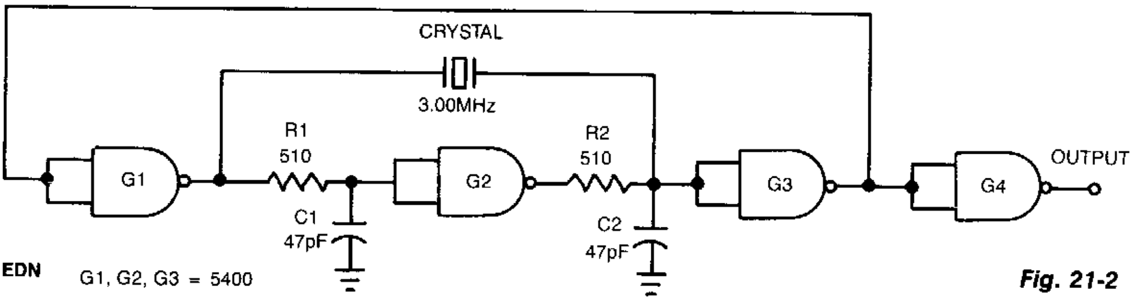


Copyright of Motorola, Inc. Used by permission.

Fig. 21-1

For frequencies below 20 MHz, a fundamental-frequency crystal can be used and the resonant tank is no longer required. Also, at this lower frequency range the typical MECL 10,000 propagation delay of 2 ns becomes small compared to the period of oscillation, and it becomes necessary to use a noninverting output. Thus, the MC10116 oscillator section functions simply as an amplifier. The 1.0 KΩ resistor biases the line receiver near V_{BB} and the 0.1-μF capacitor is a filter capacitor for the V_{BB} supply. The capacitor, in series with the crystal, provides for minor frequency adjustments. The second section of the MC10116 is connected as a Schmitt-trigger circuit; this ensures good MECL edges from a rather slow, less than 20-MHz input signal. The third stage of the MC10116 is used as a buffer and to give complementary outputs from the crystal oscillator circuit. The circuit has a maximum operating frequency of approximately 20 MHz and a minimum of approximately 1 MHz; it is intended for use with a crystal which operates in the fundamental mode of oscillation.

EASY START-UP CRYSTAL OSCILLATOR

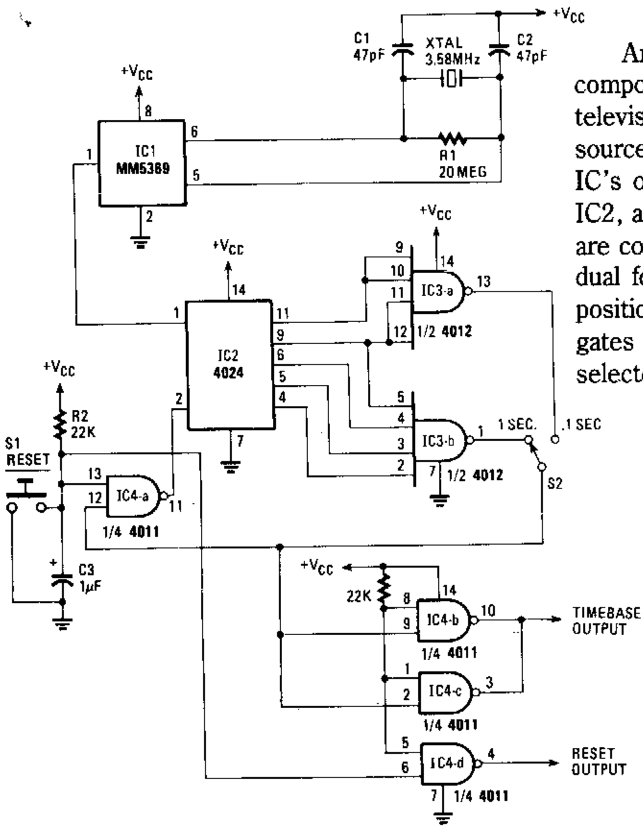


EDN

Fig. 21-2

This low cost, crystal-controlled oscillator uses one TTL gate. Two factors ensure oscillator start-up: The connection of NAND gates G1, G2, and G3 into an unstable logic configuration and the high loop gain of the three inverters. Values of R1, R2, C1, and C2 aren't critical; select them so the oscillator operates at a frequency 70 to 90% higher than the crystal frequency when the crystal is disconnected. For 1–2 MHz operation, a low-power 54L00 IC is recommended; for a 2–6 MHz, a standard 5400 type; and for 6–20 MHz, a 54H00 or 54S00.

CRYSTAL TIMEBASE



An on-board oscillator and a 17-stage divider compose IC1. By connecting a standard 3.58-MHz, television color-burst crystal as shown, an accurate source of 60-Hz squarewaves is generated at the IC's output, pin 1. Those pulses are then fed to IC2, a 4024 seven-stage ripple counter. Its outputs are connected to different gates in IC3, which is a dual four-input NAND gate. Depending on which position pulse-select switch S2 is on, one of those gates will provide an output/reset pulse of the selected width.

Fig. 21-3

LOW-FREQUENCY PIERCE OSCILLATOR

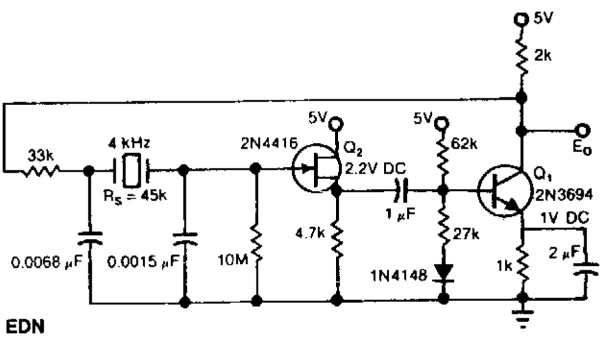


Fig. 21-4

The Pierce circuit oscillates at 4 kHz. At low frequencies, the crystal's internal series resistance R_S is quite high (45 K at 4 kHz). Therefore, an FET-based source follower is included to prevent Q1 from loading the crystal output.

1-MHz PIERCE OSCILLATORS

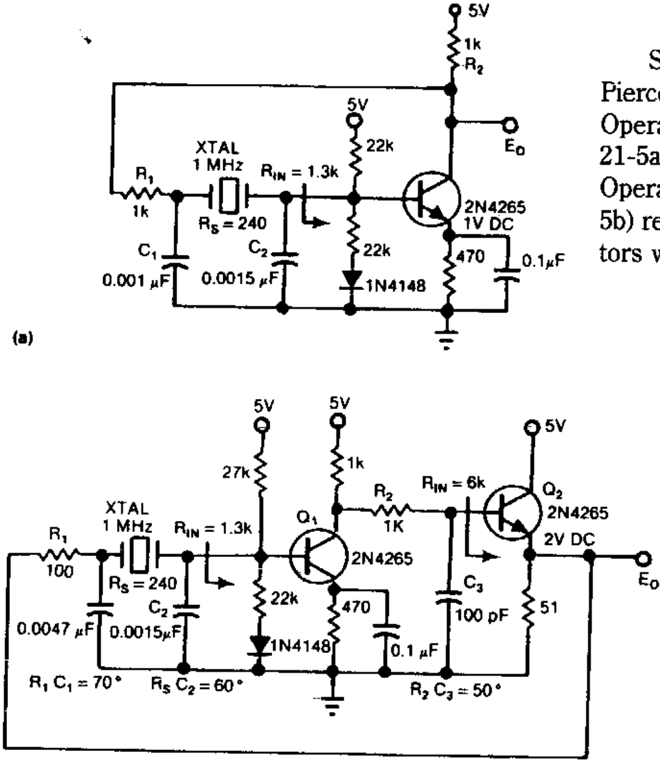


Fig. 21-5

Simple network design is a key feature of the Pierce circuit, as these 1-MHz oscillators illustrate. Operating the crystal slightly above resonance (Fig. 21-5a) requires only one high-gain transistor stage. Operating it exactly at series resonance (Fig. 21-5b) requires an extra RC phase lag and two transistors which can have lower gain.

SIMPLE CMOS CRYSTAL OSCILLATOR

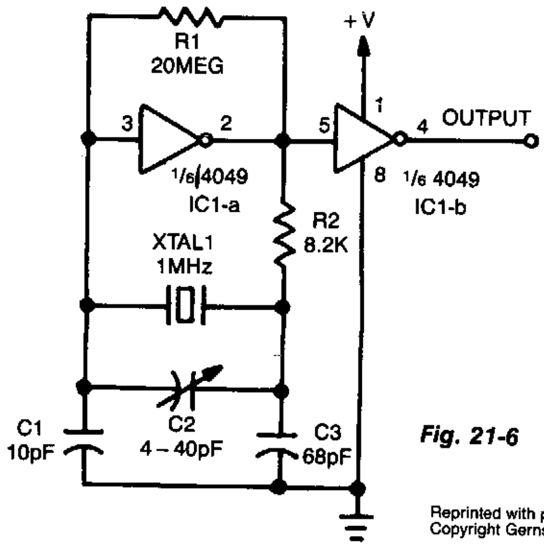
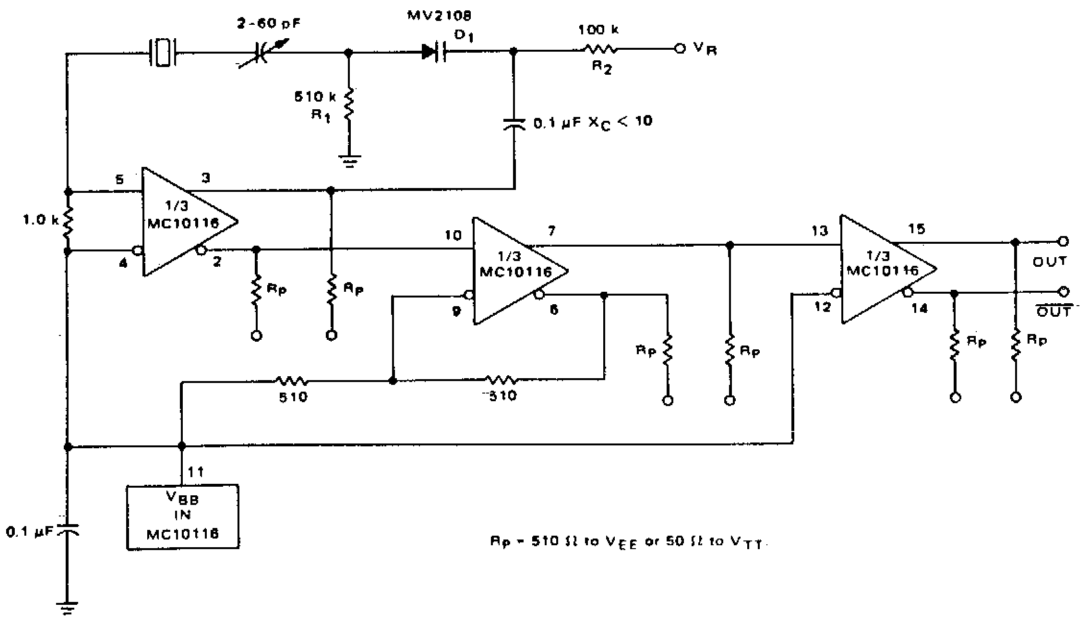


Fig. 21-6

The circuit is an inverter set up as a linear amplifier. Adding the crystal and capacitors to the feedback path, we turn the amplifier into an oscillator and force it to oscillate at, or least very near, the crystal's resonant frequency. Trimmer capacitor C2 adjusts the actual operating frequency of the circuit. The crystal should be a parallel-resonant type; maximum frequency will depend partly on supply voltage, but it should be possible to go to at least 1 MHz.

VOLTAGE-CONTROLLED CRYSTAL OSCILLATOR

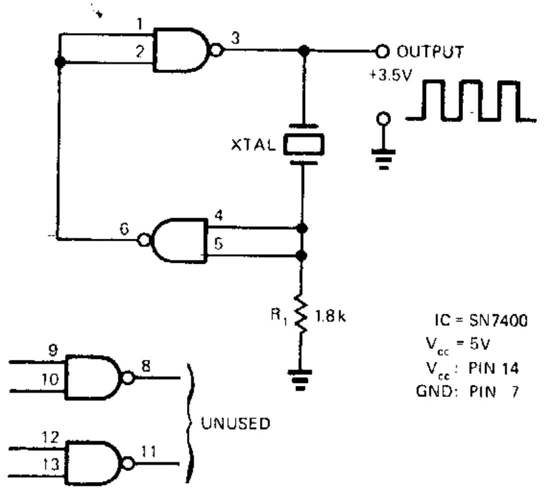


NOMINAL FREQUENCY	DEVIATION	
MHz	+PPM	-PPM
1.0000	57.0	48.0
1.8432	95.5	80.3
10.000	197.4	202.8
15.000	325.4	322.9

Fig. 21-7

A voltage-variable capacitance tuning diode is placed in series with the crystal feedback path. Changing the voltage on V_R varies the tuning diode capacitance and tunes the oscillator. The 510-KΩ resistor, R1, establishes a reference voltage for V_R—ground is used in this example. A 100-KΩ resistor, R2, isolates the tuning voltage from the feedback loop and 0.1-μF capacitor C2 provides ac coupling to the tuning diode. The circuit operates over a tuning range of 0 to 25 V. It is possible to change the tuning range from 0 to 25 V by reversing the tuning diode D1. Center frequency is set with the 2 – 60 pf trimmer capacitor. Deviation on either side of center is a function of the crystal frequency. The table in Fig. 21-7 shows measured deviation in parts per million for several tested crystals.

TWO-GATE QUARTZ OSCILLATOR

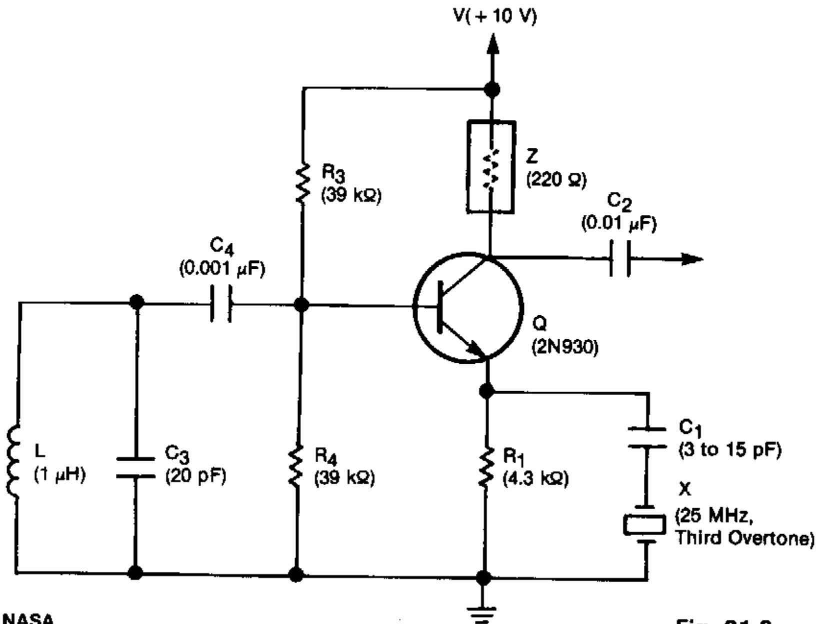


EDN

Fig. 21-8

A SN7400 quartz crystal and a resistor provide a square-wave output of approximately 3.5 V. The circuit operates reliably at frequencies from 120 kHz to 4 MHz.

CRYSTAL-CONTROLLED REFLECTION OSCILLATOR

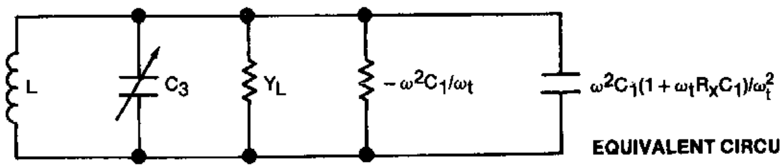


NASA

OSCILLATOR

Fig. 21-9

CRYSTAL-CONTROLLED REFLECTION OSCILLATOR (*Cont.*)

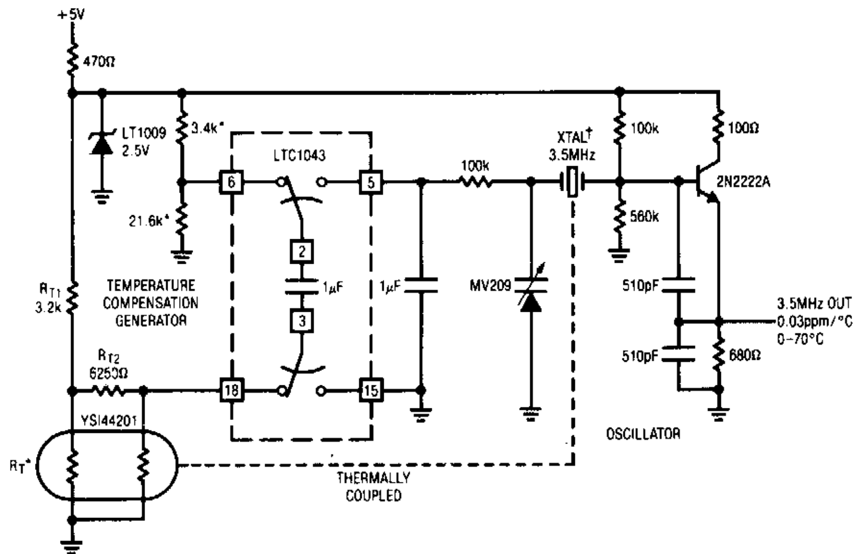


EQUIVALENT CIRCUIT

This unit is easily tunable and stable, consumes little power, and costs less than other types of oscillators that operate at the same frequencies. This unusual combination of features is made possible by a design concept that includes operation of the transistor well beyond the 3 dB frequency of its current-versus-frequency curve. The concept takes advantage of newly available crystals that resonate at frequencies up to about 1 GHz.

The emitter of transistor Q is connected with variable capacitor C1 and series-resonant crystal X. The emitter is also connected to ground through bias resistor R1. The base is connected to the parallel combination of inductor L and capacitor C3 through dc-blocking capacitor and C4 and is forward biased with respect to the emitter by resistors R3 and R4. Impedance Z could be the 220-Ω resistor shown or any small impedance that enables the extraction of the output signal through coupling capacitor C2. If Z is a tuned circuit, it is tuned to the frequency of the crystal.

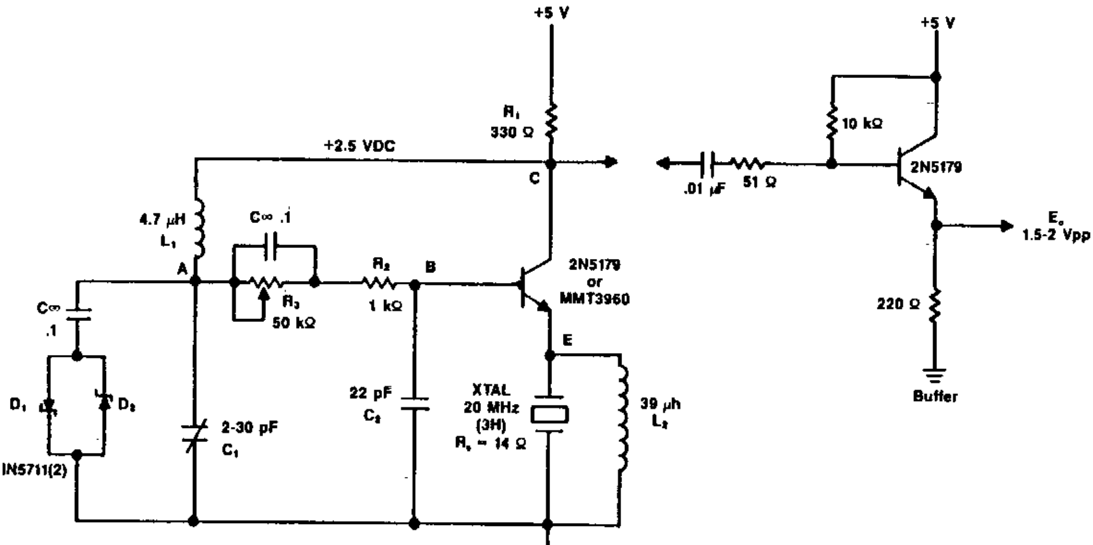
TEMPERATURE-COMPENSATED CRYSTAL OSCILLATOR



LINEAR TECHNOLOGY CORP. *Fig. 21-10*

This circuit uses LTC1043 to differentiate between a temperature sensing network and a dc reference. The single-ended output biases a varactor-tuned crystal oscillator to compensate drift. The varactor crystal network has high dc impedance, eliminating the need for an LTC1043 output amplifier.

20-MHz VHF CRYSTAL OSCILLATOR

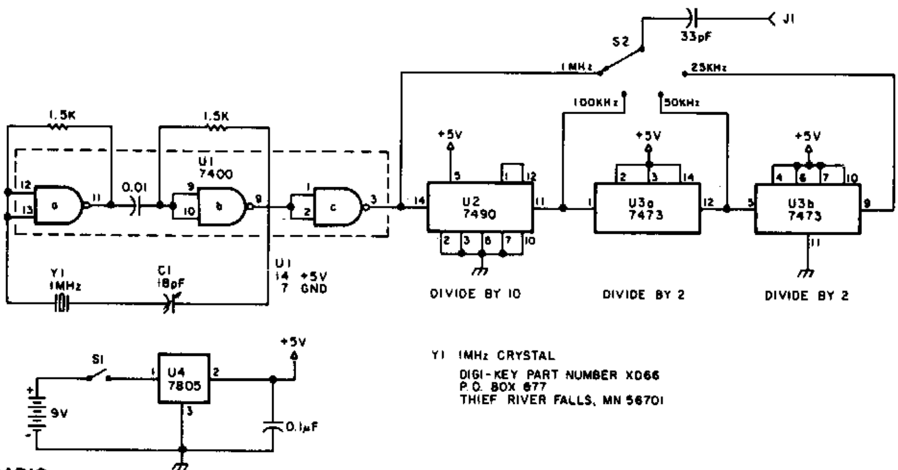


R. Matthys, RF Design, March 1987, p 31.

Fig. 21-11

A typical circuit at 20 MHz is shown. The crystal, which has an internal series resistance R_S of 14 Ω, oscillates at its third harmonic. The diode clamp D1 and D2 provides a constant amplitude control. The transistor operates continuously in a linear mode over a complete cycle of oscillation, and reflects a reasonably constant load across the crystal at all times.

MARKER GENERATOR

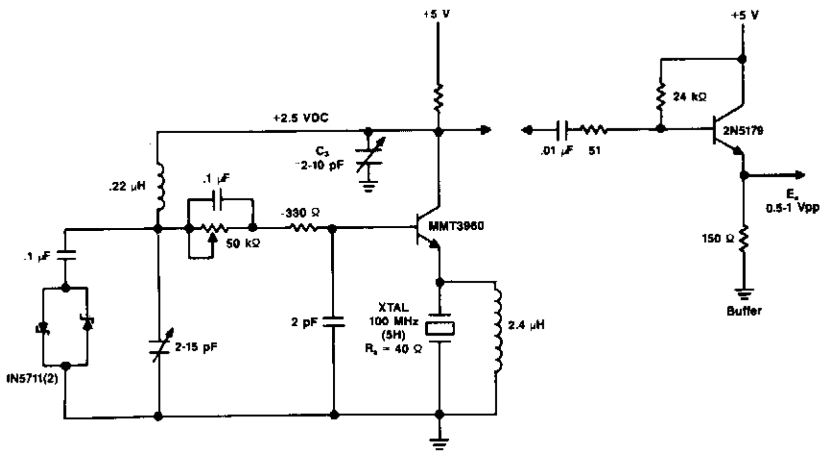


73 AMATEUR RADIO

Fig. 21-12

The oscillator section uses three sections of a 7400 quad NAND gate integrated circuit. The 1-MHz signal from the oscillator is fed into a 7490 decade counter configured to divide by ten, providing the 100-kHz signal. To obtain the 50 and 25 kHz outputs, the 100-kHz signal is further divided by 7473 dual J-K flip-flop. The first half of the 7473 divides the 100-kHz signal by two, yielding the 50 kHz signal. The second half of the 7473 again divides by two, giving the 25 kHz signal. S2 selects the output, a square wave, rich in harmonics. The generator can be powered from any convenient 6 to 12 Vdc source. A 7805 fixed-voltage regulator supplies the regulated voltage for the oscillator and the divider chips. The generator described here is powered by a 9-V transistor radio battery.

100-MHz VHF CRYSTAL OSCILLATOR

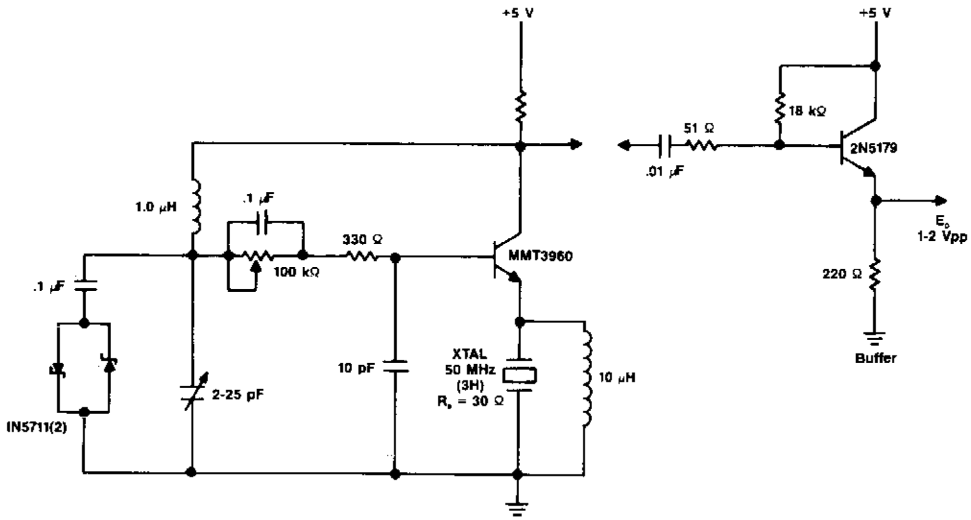


R. Matthys, RF Design, March 1987, p. 31.

Fig. 21-13

Figure 21-13 shows a 100-MHz oscillator operating on the fifth harmonic. Again to maintain the transistor's gain, note the increase in the collector's load resistance R1 because of the increase in the quartz crystal's internal series resistance R_S. C3 is needed at frequencies above 50 MHz to tune out the shunting effect of L1 on R1, to maintain a high load resistance for the transistor and get enough gain for oscillation. The equivalent series $R_L C_L$ load across the crystal is 8.2 Ω (R_L) and 200 pF (C_L).

50-MHz VHF CRYSTAL OSCILLATOR



R. Matthys, RF Design, March 1987, p. 31.

Fig. 21-14

Figure 21-14 shows a 50-MHz oscillator operating on a third harmonic. The collector's load resistor R1 has been increased because the quartz crystal's internal series resistance R_S increases with frequency in the VHF range. The crystal's internal series resistance R_S is 30 Ω, and the transistor's minimum current gain H_{FE} is 100. Using the same technique as for the 20 MHz oscillator, the external series $R_L C_L$ equivalent load seen by the 50 MHz crystal is 5.6 Ω (R_L) and 1000 pF (C_L).

22

Decoders

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Second-Audio Program Adapter
Tone Decoder
Encoder/Decoder
Direction Detector/Decoder
Sound-Activated Decoder

SECOND-AUDIO PROGRAM ADAPTER

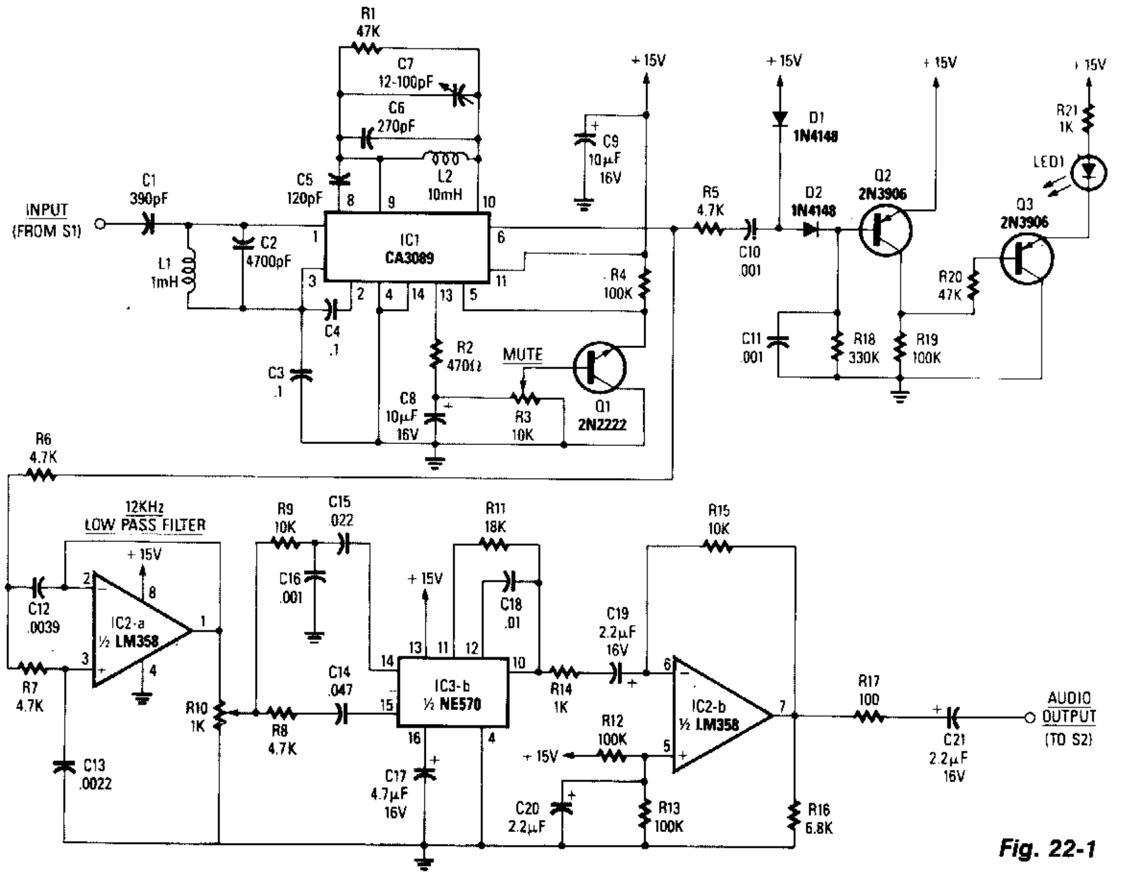


Fig. 22-1

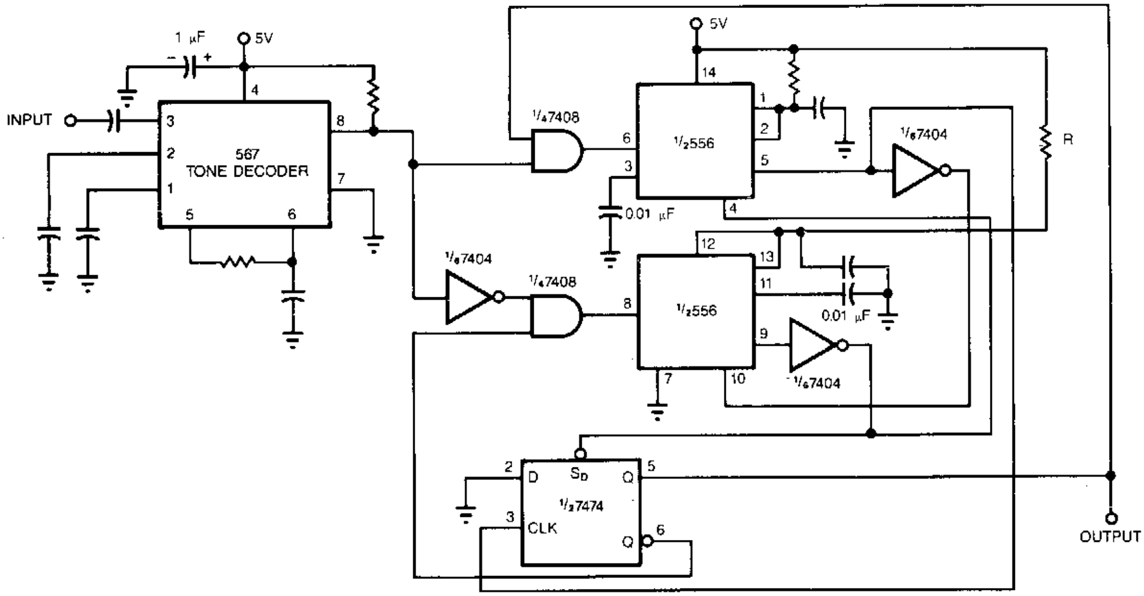
The baseband-audio input comes from the pole of switch S1 in the stereo decoder, and is coupled to IC1 (a CA3089) via a 78.6 kHz bandpass filter that consists of capacitors C1 and C2, and inductor L1. IC1 is a combination i-f amplifier and quadrature detector normally used for FM radio systems operating within an i-f of 10.7 MHz. The device works equally well at 78.6 kHz. Capacitors C6 and C7, and inductor L2 tune the detector section to 78.6 kHz, while C5 provides the necessary 90-degree phase shift for proper quadrature detector operation. The output voltage at pin 13 of IC1 is proportional to the level of the incoming signal. When the voltage at the wiper of potentiometer R3 reaches a predetermined threshold level, Q1 conducts, grounding pin 5 of IC1, enabling IC1's mute function.

Detected audio output from pin 6 of IC1 goes to IC2a, which is configured as a 12-kHz, −12 dB per octave, low-pass filter. The output of IC2a appears across potentiometer R10, which provides a means of adjusting the drive level into IC3b, the 2:1 compander.

Audio from the wiper of R10 is split into two paths: a high-pass filter (C14 and R8) provides a path to the rectifier input of the compander, and a bandpass filter (R9, C16, and C15) that feeds the audio input of the compander. A fixed 390-μs de-emphasis network is formed by C18 and R11 in conjunction with IC3b. Corrected audio appears at pin 10 of IC3b and is coupled to IC2b, and output buffer amplifier.

Audio from pin 6 of IC1 is also coupled to an audio high-pass filter, R5 and C10, and fed to an audio rectifier, D1, D2, and C11. When a SAP signal is detected by IC1, it is rectified by D1 and D2; the resultant dc charges C11. An increasing positive voltage at the base of Q2 causes its current flow to decrease, so the voltage at Q2's collector also decreases. That in turn causes the base voltage of Q3 to drop, which causes Q3 to conduct, thereby lighting the LED.

TONE DECODER

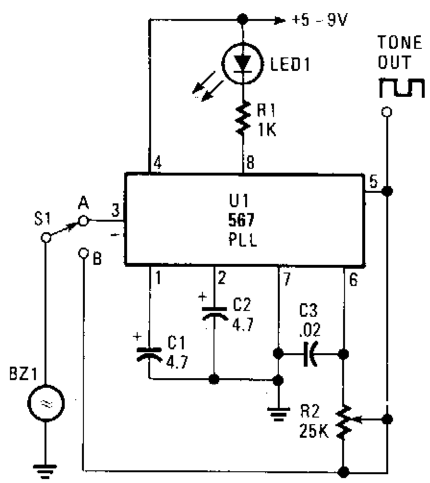


HANDS-ON ELECTRONICS/POPULAR ELECTRONICS *Fig. 22-2*

Adding a pair of one shots to the output of a 567 tone decoder renders it less sensitive to out-of-band signals and noise. Without the one shots, the 567 is prone to spurious output chatter. Other protection schemes, such as feeding back outputs or using an input filter, do not work as well as the one shots. The output of the 567 is high in the absence of a tone and becomes low when it detects a tone. The tone decoder triggers a one shot via an AND gate. The one shot's period is set to slightly less than the duration of a tone burst.

When the output of the tone decoder decreases, it triggers the second one shot. The second one shot's period is set to slightly less than the interval between tone bursts. The flip-flop enables and disables the inputs to one shots so that spurious outputs from the tone decoder do not affect the output.

ENCODER/DECODER



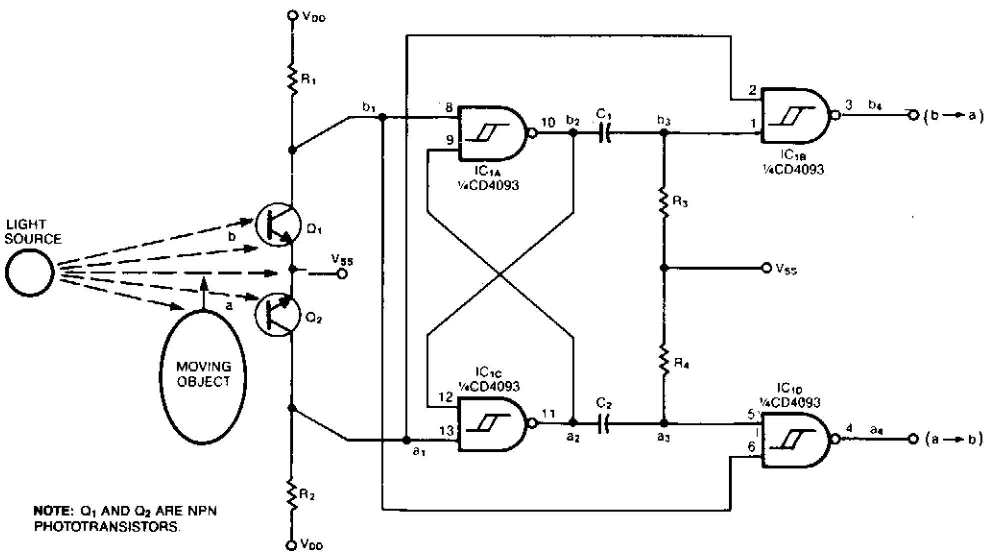
EDN

Fig. 22-3

The transducer circuit can be operated as either a tone encoder or decoder by changing the position of S1. The operating frequency of that dual-purpose circuit is determined by C3 and R2. Capacitors C1 and C2 are not critical and can be of almost any value between 1 and 5 mf. When the circuit is receiving an on-frequency signal, LED1 lights. Although a two wire piezo transducer with a resonance frequency of 2500 Hz was used in the circuit, any piezo unit should work—as long as the values of C3 and R2 are selected to tune to the transducer's operating frequency.

With power on and S1 in the B position, adjust R2 for the loudest tone output. The circuit should be tuned to the resonance frequency of the transducer. In that position, the circuit can be used as an acoustical or tone signal encoder. Next, switch to the A position and aim an on-frequency audible tone toward the transducer; the LED should light, indicating a decoded signal.

DIRECTION DETECTOR DECODER



HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

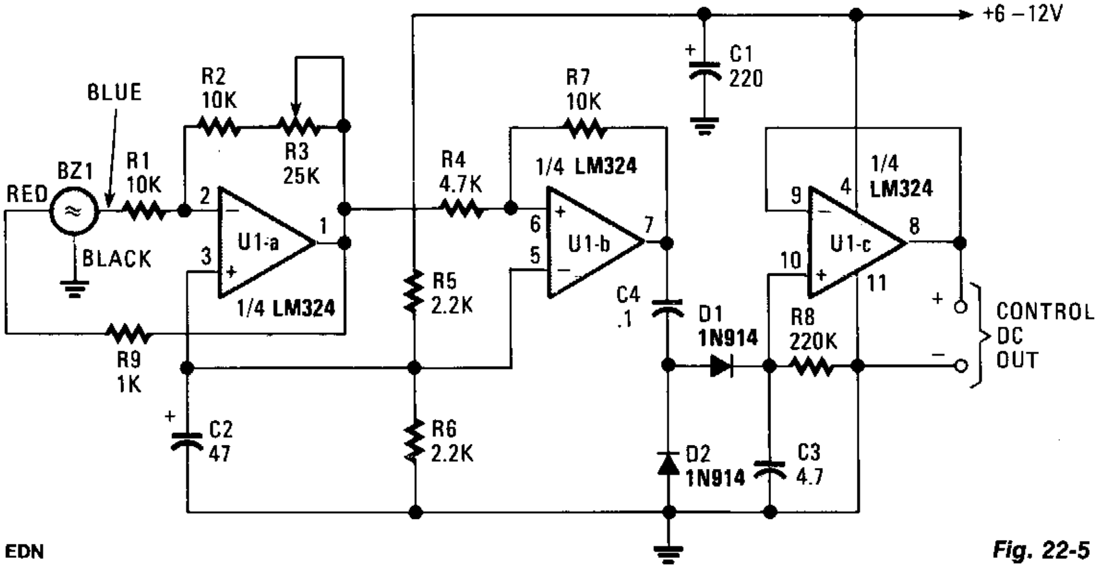
Fig. 22-4

This circuit, which was developed to monitor the traffic of bumblebees in and out of the hive, differentiates a-to-b motion from b-to-b motion. When used with an optical decoder, the circuit distinguishes clockwise from counterclockwise rotation and provides a resolution of one output pulse per quadrature cycle.

Q1 and Q2 are mounted so that a moving object first blocks one phototransistor, then both, then the other. Depending on the direction in which the object is moving, either IC1B or IC1D emits a negative pulse when the moving object blocks the second sensor. An object can get as far as condition 3 and retreat without producing an output pulse; that is, the circuit ignores any probing or jittery motion. If an object gets as far as condition 4, however, a retreat will produce an opposite-direction pulse.

The time constants R3C1 and R4C2 set the output pulse width. A 100 KΩ/100pF combination, for example, produces 10-μs pulses. Select a value for pullup resistors R1 and R2 from the 10 K to 100 KΩ range, according to the sensitivity your application requires.

SOUND-ACTIVATED DECODER



EDN *Fig. 22-5*

The piezo transducer operates as a sound-pickup device as well as a frequency-selective filter. By controlling the gain of the op amps, the oscillator can be transformed into a sensitive and frequency-selective tone-decoder circuit. With the gain of U1a set just below the point of self oscillation, the transducer becomes sensitive to acoustically coupled audio tones that occur at or near its resonant frequency.

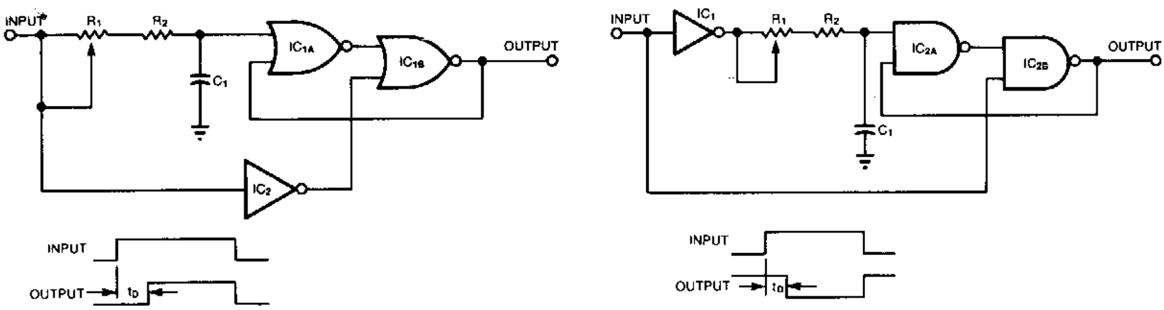
The circuit's output can be used to activate optocouplers, drive relay circuits, or to control almost any dc-operated circuit. The dc signal at the output of U1c varies with 0 to over 6 V, depending on the input-signal level. One unusual application for the sound-activated decoder would be in extremely high-noise environments, where normal broadband microphone pickup would be useless. Because piezo transducers respond only to frequencies within a very narrow bandwidth, little if any of the noise would get through the transducer.

23

Delay Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

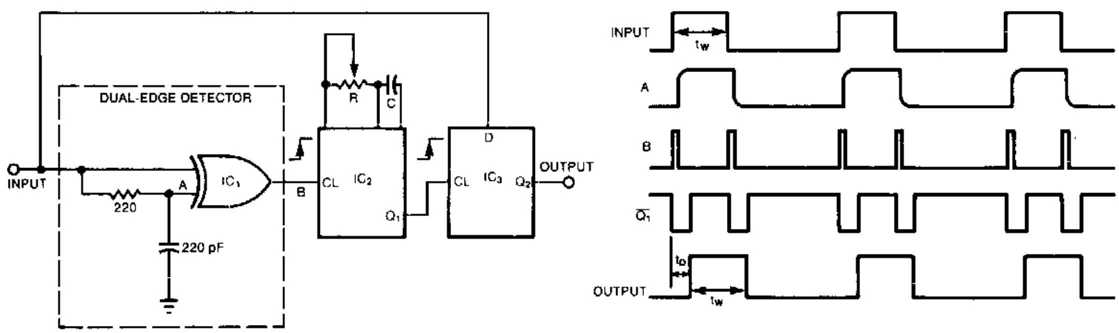
LEADING-EDGE DELAY



EDN (a) (b) *Fig. 23-1*

Circuit (a) lets you delay the leading edge of a positive pulse while leaving the trailing edge almost unaffected. A positive input transition, inverted by IC2, has no effect on IC1B. However, when the positive transition reaches IC1A, (delayed by the adjustable network of R1, R2, and C1), it toggles both NOR gates, initiating the output pulse. When the input decreases IC1B follows suit, delayed only by the propagation through itself and IC2. Circuit (b) produces an inverted output pulse. Inverter IC1 serves as a buffer for the signal source—an advantage when driving a low-impedance (short-delay) network. Moreover, only the propagation delay of IC2B separates the output's trailing edge from that of the input. You can configure circuit (a) to handle negative pulses by using NAND instead of NOR gates. Similarly, circuit (b) will produce a delayed positive pulse in response to negative input pulse, if you substitute NOR gates for NAND gates.

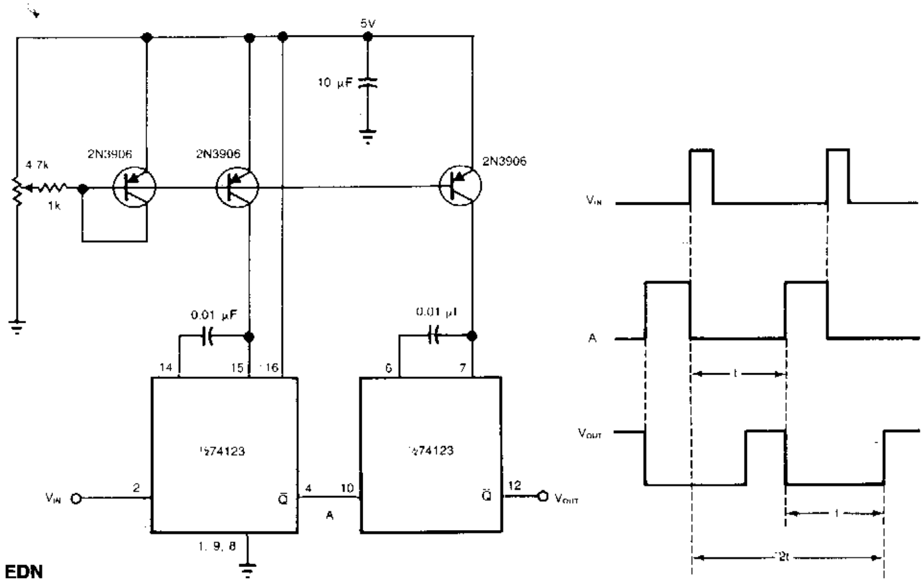
PULSE DELAY WITH DUAL-EDGE TRIGGER



EDN *Fig. 23-2*

A single monostable multivibrator delays a pulse train by a variable amount; nonetheless, this amount can be no less than the minimum allowed pulse width t_w. The exclusive-OR gate, IC1, generates a short pulse following every leading or falling edge of the input waveform. These pulses cause one-shot IC2 to produce a negative-going pulse with a duration equal to the desired time delay t_p, which you set by adjusting potentiometer R. Flip-flop IC3 then creates a delayed replica of the input pulse by latching the Q1 output of IC2 between positive-going transitions. You can independently control the output-pulse duration by cascading a second one shot with the first.

ADJUSTABLE DELAY



EDN

Fig. 23-3

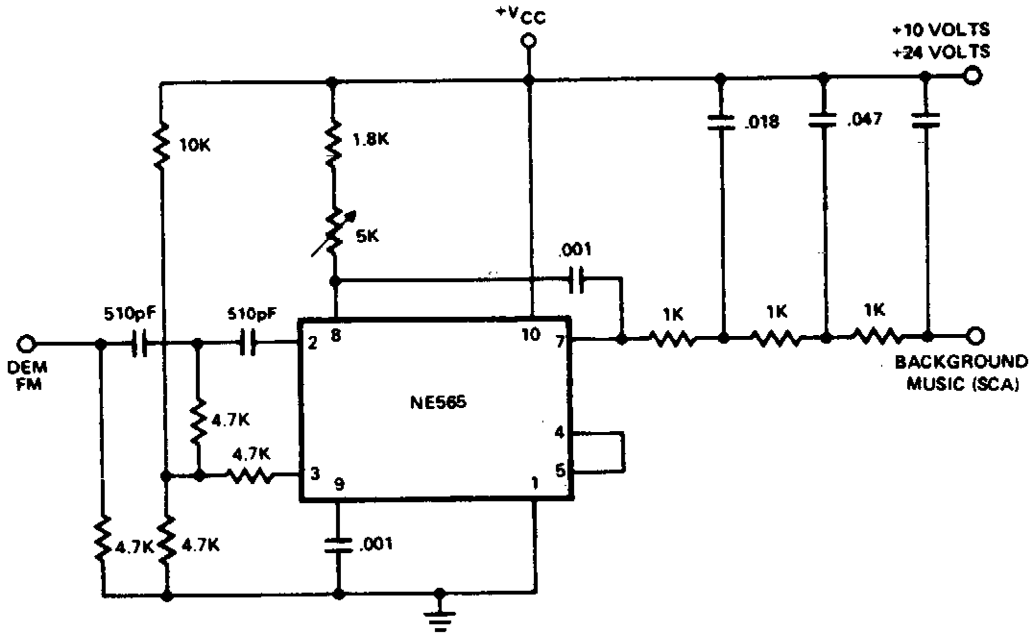
You can obtain well over 360° of phase delay by cascading two monostable multivibrators. In a typical configuration, a single monostable multivibrator is used to introduce delay in a pulse train; the multivibrator triggers on each incoming pulse, provided it resets in time for the next pulse. Yet even when it resets in time, the single monostable multivibrator provides a maximum phase delay of less than 360°. However, with the cascaded-multivibrator approach, you can achieve 650° of phase delay by using an input-pulse spacing of 200 μs for example, with the component values shown. Every input pulse will trigger the circuit while you adjust the phase delay throughout its available range. The first multivibrator triggers the second one, whose reset marks the total delay time (2*t*). Each introduces a delay of *t* μs, based on 0.01-μF timing capacitors and equal charging currents from the three-transistor, dual-current source. The two multivibrator arrangement allows the first multivibrator to reset in time to be triggered by the next input pulse. Also, the variation of *t* is linear with the potentiometer setting.

24

Demodulator

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

565 SCA DEMODULATOR



SIGNETICS

Fig. 24-1

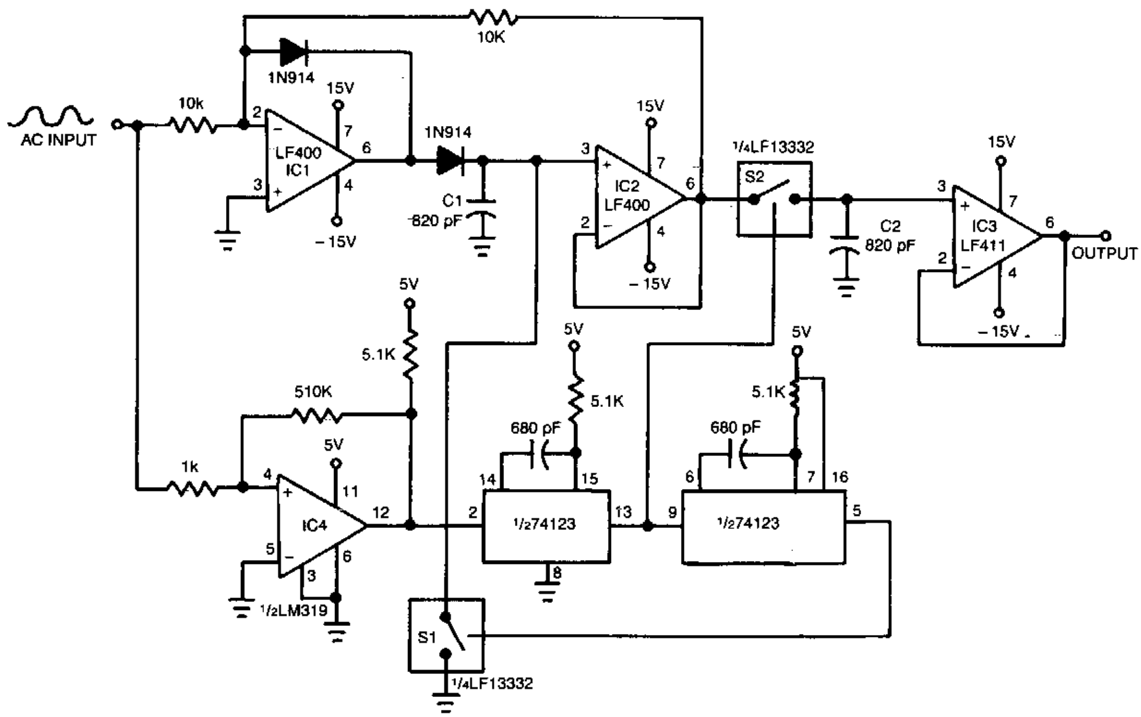
This application involves demodulation of a frequency-modulated subcarrier of the main channel. This popular example uses the PLL to recover the SCA (Subsidiary Carrier Authorization or storecast music) signal from the combined signal of many commercial FM broadcast stations. The SCA signal is a 67 kHz frequency-modulated subcarrier which puts it above the frequency spectrum of the normal stereo or monaural FM program material. By connecting the circuit to a point between the FM discriminator and the deemphasis filter of an FM receiver and tuning the receiver to a station which broadcasts an SCA signal, you can obtain hours of commercial-free background music.

25

Detectors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

WIDE-RANGE PEAK DETECTOR

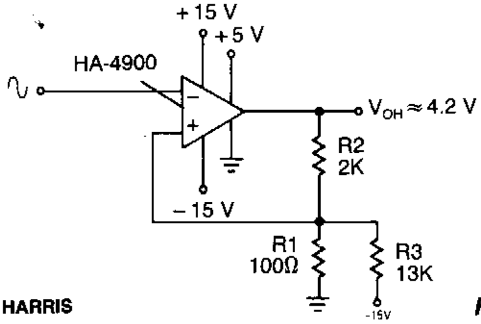


EDN

Fig. 25-1

IC1 and IC2 form an inverting half-wave precision-rectifier/peak-detector circuit. Negative input-signal, swings with peaks larger than the voltage on C1, cause this capacitor to charge to the new peak voltage. The capacitor holds this voltage until a larger signal peak arrives. When the input swings high, comparator IC4 detects the zero crossing and triggers the one-shot multivibrator. The one shot closes FET switch S2, thereby causing C2 to charge to the peak voltage held on C1, during the previous half cycle. The second one shot then produces a pulse that causes FET switch S1 to discharge C1. If the next negative signal-input peak is different from the previous one, the circuit captures it and it appears at IC3's output during the next half cycle. The peak detector thus resets itself once every input-waveform cycle. Note that the zero crossings are necessary to trigger the switches; therefore, the circuit is usable only with ac signals.

SCHMITT TRIGGER



HARRIS

Fig. 25-2

This circuit has a 100-mV hysteresis which can be used in applications where very fast transition times are required at the output, even though the signal input is very slow. The hysteresis loop also reduces false triggering because of noise in the input.

ANALOG PEAK DETECTOR WITH DIGITAL HOLD

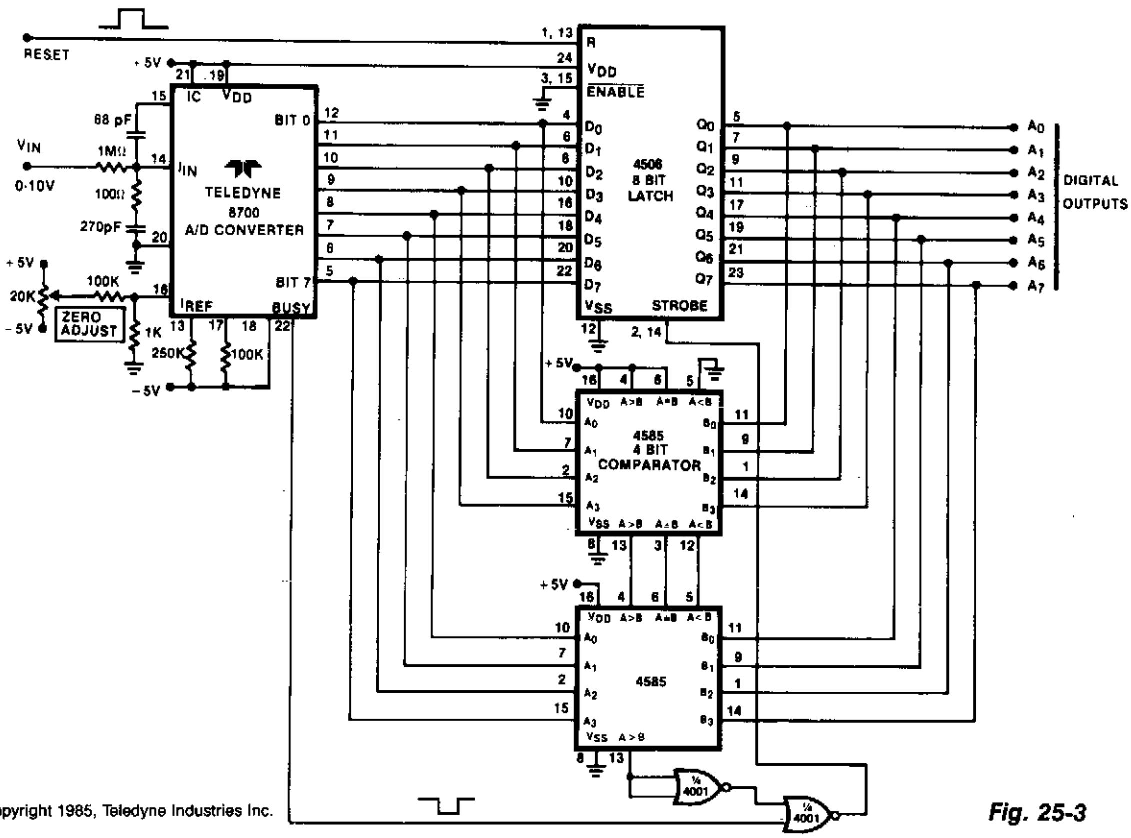
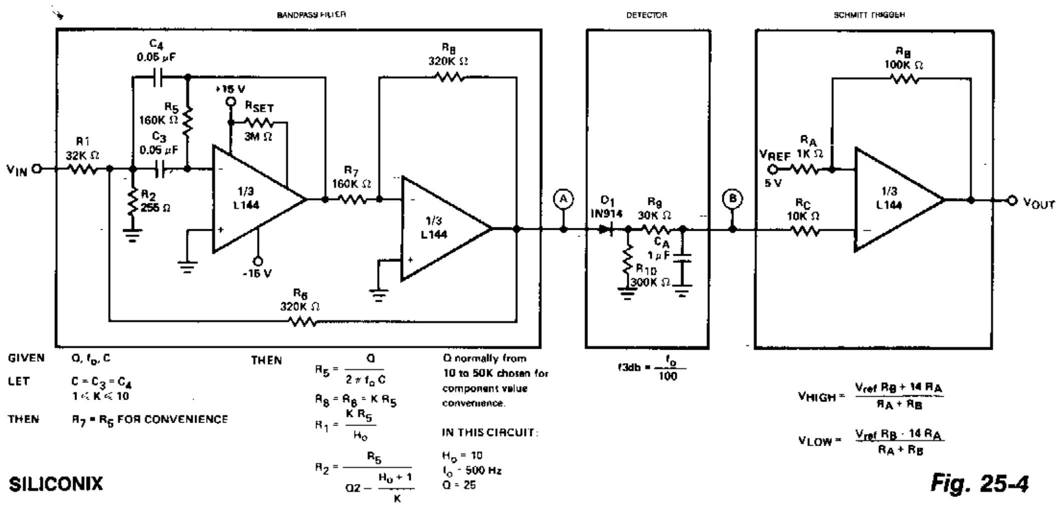


Fig. 25-3

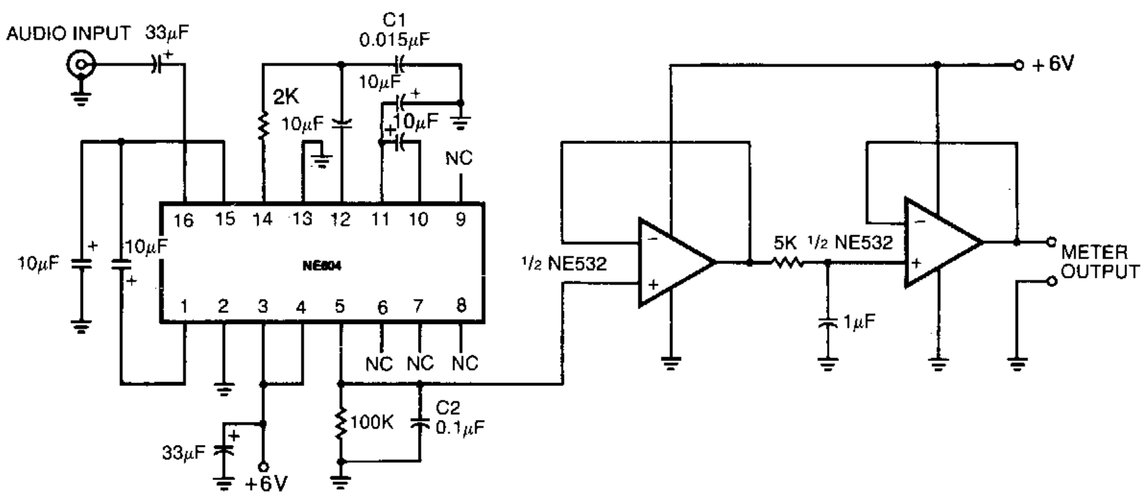
Analog peak detection is accomplished by repeatedly measuring the input signal with an a/d converter and comparing the current reading with the previous reading. If the current reading is larger than the previous, the current reading is stored in the latch and becomes the new peak value. Since the peak is stored in a CMOS latch, the peak can be stored indefinitely.

500-Hz TONE DETECTOR



SILICONIX *Fig. 25-4*

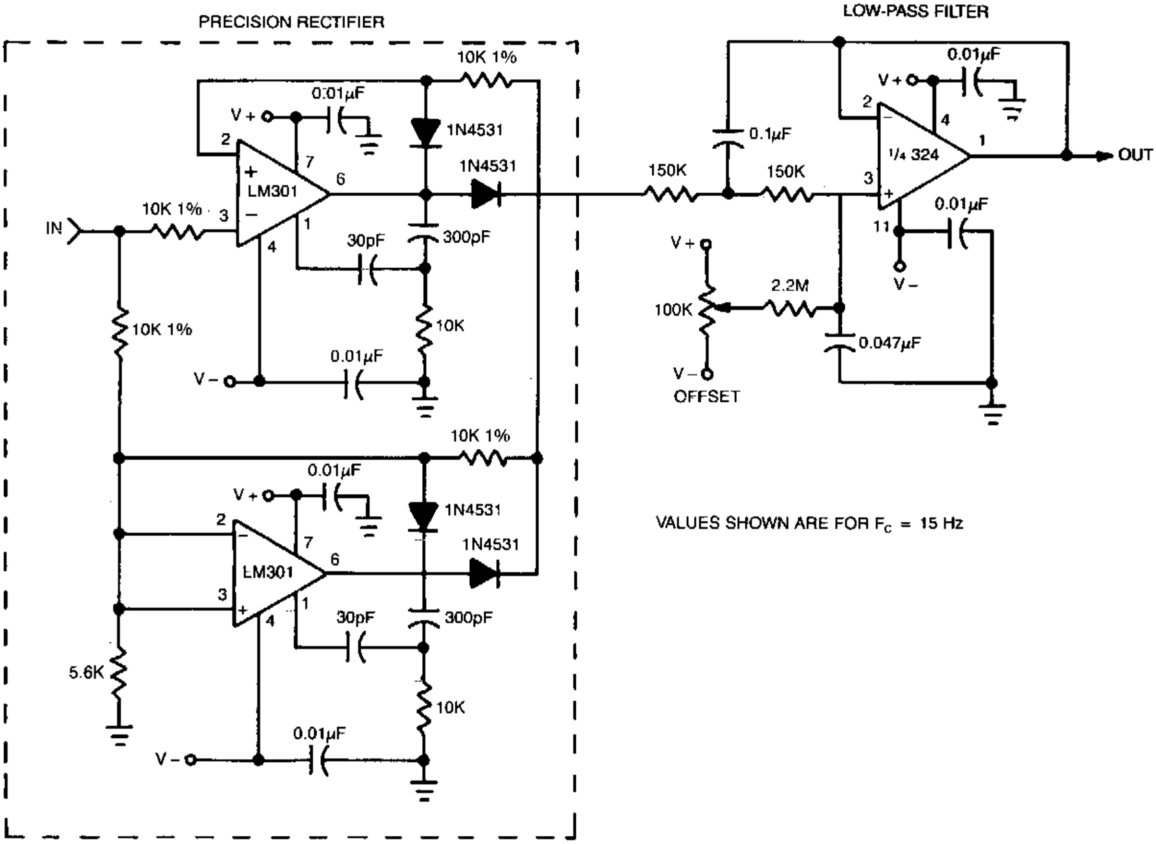
AUDIO DECIBEL LEVEL DETECTOR WITH METER DRIVER



SIGNETICS *Fig. 25-5*

This circuit draws very little power, less than 5 mA with a single 6-V power supply, making it ideal for portable battery-operated equipment. The small size and low power consumption belie the 90-dB dynamic range and 10.5-μV sensitivity. Dc output voltage proportional to the $\log_{10}$ of the input signal level. Thus, a standard 0 – 5 voltmeter can be linearly calibrated in decibels over a single 80-dB range. The circuit is within 1.5-dB tolerance over the 80-dB range for audio frequencies from 100 Hz to 10 kHz. Higher audio levels can be measured by placing an attenuator ahead of the input capacitor.

PRECISION ENVELOPE DETECTOR



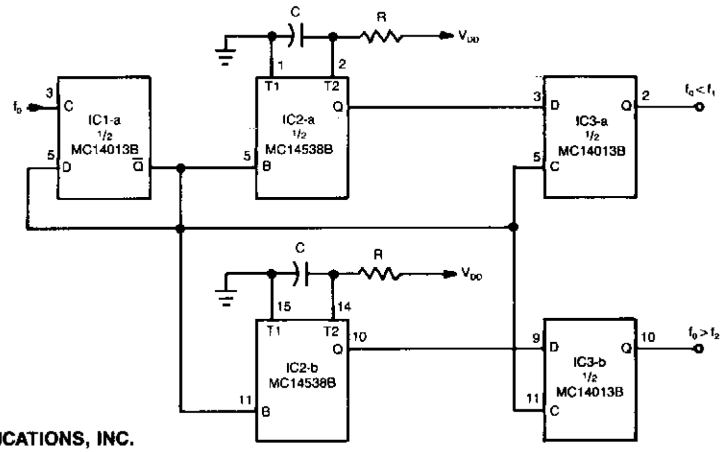
EDN

Fig. 25-6

This circuit is useful for signal-processing sonar data recorded on an instrumentation-quality analog tape recorder. The envelope detector utilizes ready available parts, and furnishes accuracy beyond 100 kHz. Two LM301 op amps connected as precision absolute-value circuits use 2-pole frequency compensation for increased slew rate. And one section of an LM324 quad op amp connected in a Butterworth LPF configuration subjects the rectifier's output to a low-pass filter.

$$f_c = 1/2\pi RC_1$$
$$C_2 = (1/2)\ C_1$$

FREQUENCY-BOUNDARY DETECTOR

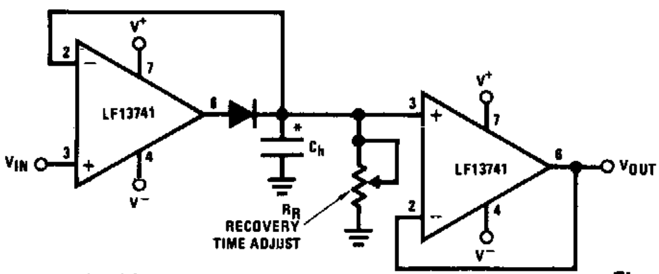


GERNSBACK PUBLICATIONS, INC. *Fig. 25-7*

The circuit can be used to tell whether or not an input signal is within a certain frequency range. The device consists of three ICs, a dual monostable multivibrator, and two dual D-type flip-flops. The signal whose frequency is in question is fed to the clock input of one of the flip-flops. The *Q* output of that flip-flop (IC1a) is cross coupled to its data input so that it acts like a divide-by-two counter. The trailing edge of the *Q* output is used to trigger the one shots formed by IC2. The upper- and lower-frequency boundaries are determined by the two sections of IC2; the dual precision monostable multivibrator and their external rc networks. The upper-frequency boundary, f1, is set by the output of IC2a, and the lower-frequency boundary, f2, is set by the output of IC2b.

The frequency of the input to the circuit can be anywhere from dc to 100 kHz. The states of the outputs of IC2, which determine the upper- and lower-frequency boundaries, are latched by IC3a and IC3b respectively. The output of IC3a will be high only when the input frequency is less than that of the output of IC2a, f_1. The output of IC3b will be high only when the frequency of the input is greater than that of the output of IC2b, f_2.

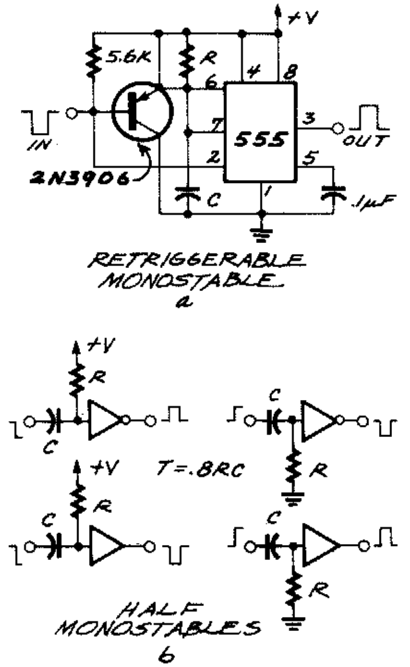
LOW-DRIFT PEAK DETECTOR



NATIONAL SEMICONDUCTOR CORP. *Low leakage capacitor *Fig. 25-8*

This circuit uses op amp U1 to compensate for the offset in peak detector diode D1. Across C_h is the exact peak voltage; U2 is used as a voltage follower to read this voltage.

EDGE DETECTOR

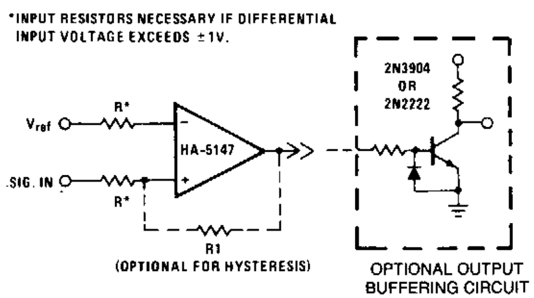


The 555 is a monostable that *wants* a negative-going trigger. If the pulse you're feeding it with is positive-going, you can run it through an inverter made up of either an inverting gate or, if you're tight on space, a single transistor. Both ways are shown. The circuits shown in Fig. 25-9b are edge detectors as well, and are usually referred to as half monostables, since they can't be used in every application. The width of the output pulse is determined by the RC value, but there are a few rules governing their use:

- The input pulse has to be wider than the output pulse
- The input pulse can't be glitchy
- The circuit can't be retriggered faster than the RC time

Fig. 25-9

PRECISION THRESHOLD DETECTOR

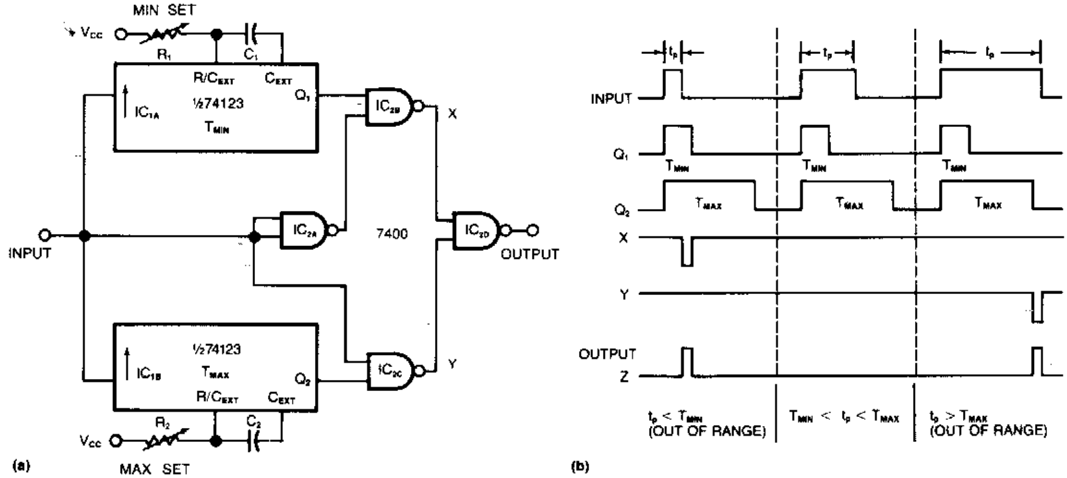


This circuit requires low noise, low and stable offset voltages, high open loop gain, and high speed. These requirements are met by the HA-5147. The standard variations of this circuit can easily be implemented using the HA-5147. For example, hysteresis can be generated by adding R1 to provide small amounts of positive feedback. The circuit becomes a pulse width modulator if V_{ref} and the input signal are left to vary. Although the output drive capability of this device is excellent, the optional buffering circuit can be used to drive heavier loads, preventing loading effects on the amplifier.

HARRIS

Fig. 25-10

OUT-OF-BOUNDS PULSE-WIDTH DETECTOR

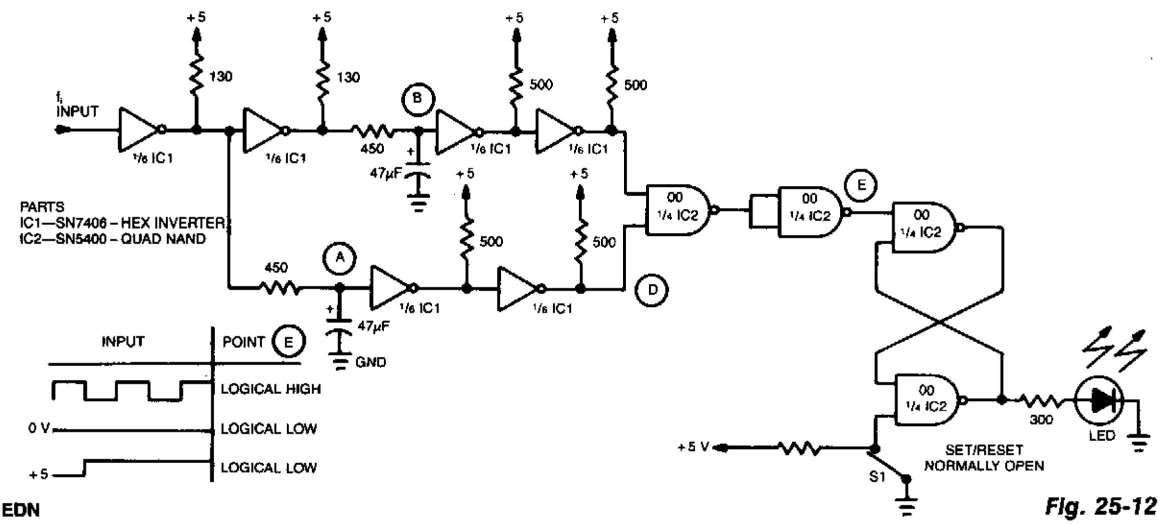


EDN

Fig. 25-11

Requiring only two ICs to monitor a train of positive pulses, this circuit produces a single positive output pulse for each input pulse whose duration is either too long or too short. You specify the minimum and maximum limits by adjusting the trimming potentiometers, R1 and R2. You can set the value of the acceptable pulse width from approximately 50 ns to 10 μs, for a 74123 monostable multivibrator. The leading edge of an input pulse triggers one shots IC1A and IC1B as you can see from the timing diagram. Each NAND-gate output is high unless either or both inputs are low, so outputs X and Y are high unless the circuit encounters an out-of-range pulse. IC2D then gates a negative pulse from IC2B or IC2C to produce the circuit's positive output pulse.

DIGITAL FREQUENCY DETECTOR



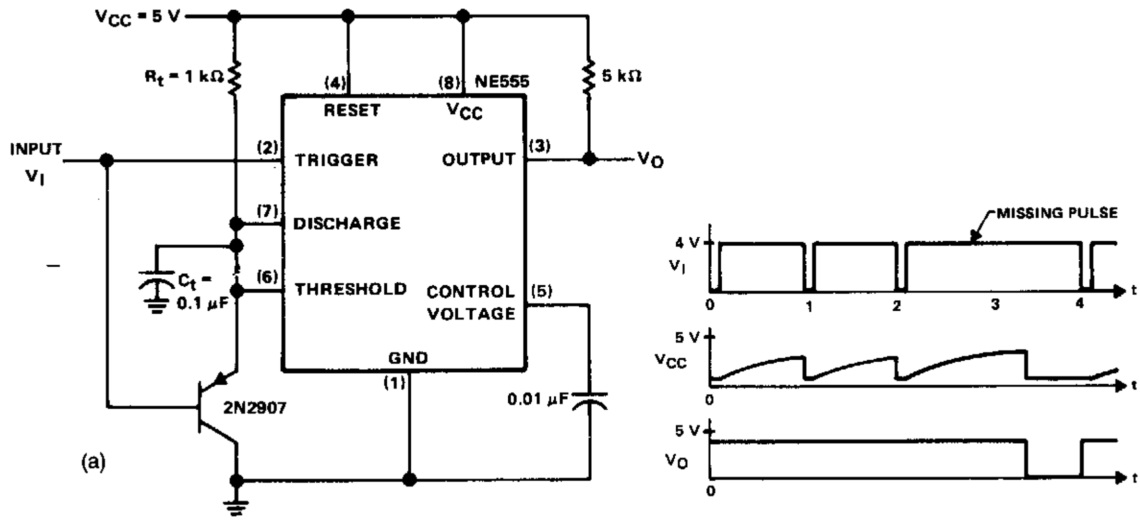
EDN

Fig. 25-12

DIGITAL FREQUENCY DETECTOR (*Cont.*)

A simple inventer and NAND gate can be connected to yield a highly compact and reliable digital frequency detector. This circuit can detect frequencies up to 3 MHz with 50% duty cycles. When a frequency, f_i, appears at the input, points A and B detect a logical high dc level. Thereupon point E increases the latch sets and the LED lights. If the input frequency is absent and if the voltage is either at a constant high or low level, points A and B will be complementary and point E will decrease. This will reset the latch and extinguish the LED.

MISSING-PULSE DETECTOR



Reprinted by permission of Texas Instruments.

(b)

Fig. 25-13

This circuit will detect a missing pulse or abnormally long spacing between consecutive pulses in a train of pulses. The timer is connected in the monostable mode. The time delay should be set slightly longer than the timing of the input pulses. The timing interval of the monostable circuit is continuously retriggered by the input pulse train, V_I. The pulse spacing is less than the timing interval, which prevents V_C from rising high enough to end the timing cycle. A longer pulse spacing, a missing pulse, or a terminated pulse train will permit the timing interval to be completed. This will generate an output pulse, V_O as illustrated in Fig. 25-3b. The output remains high on pin 3 until a missing pulse is detected at which time the output decreases.

The NE555 monostable circuit should be running slightly slower, lower in frequency, than the frequency to be analyzed. Also, the input cannot be more than twice this free-running frequency or it would retrigger before the timeout and the output would remain in the low state continuously. The circuit operates in the monostable mode at about 8 kHz, so pulse trains of 8 to 16 kHz can be observed.

DIGITAL PEAK DETECTOR

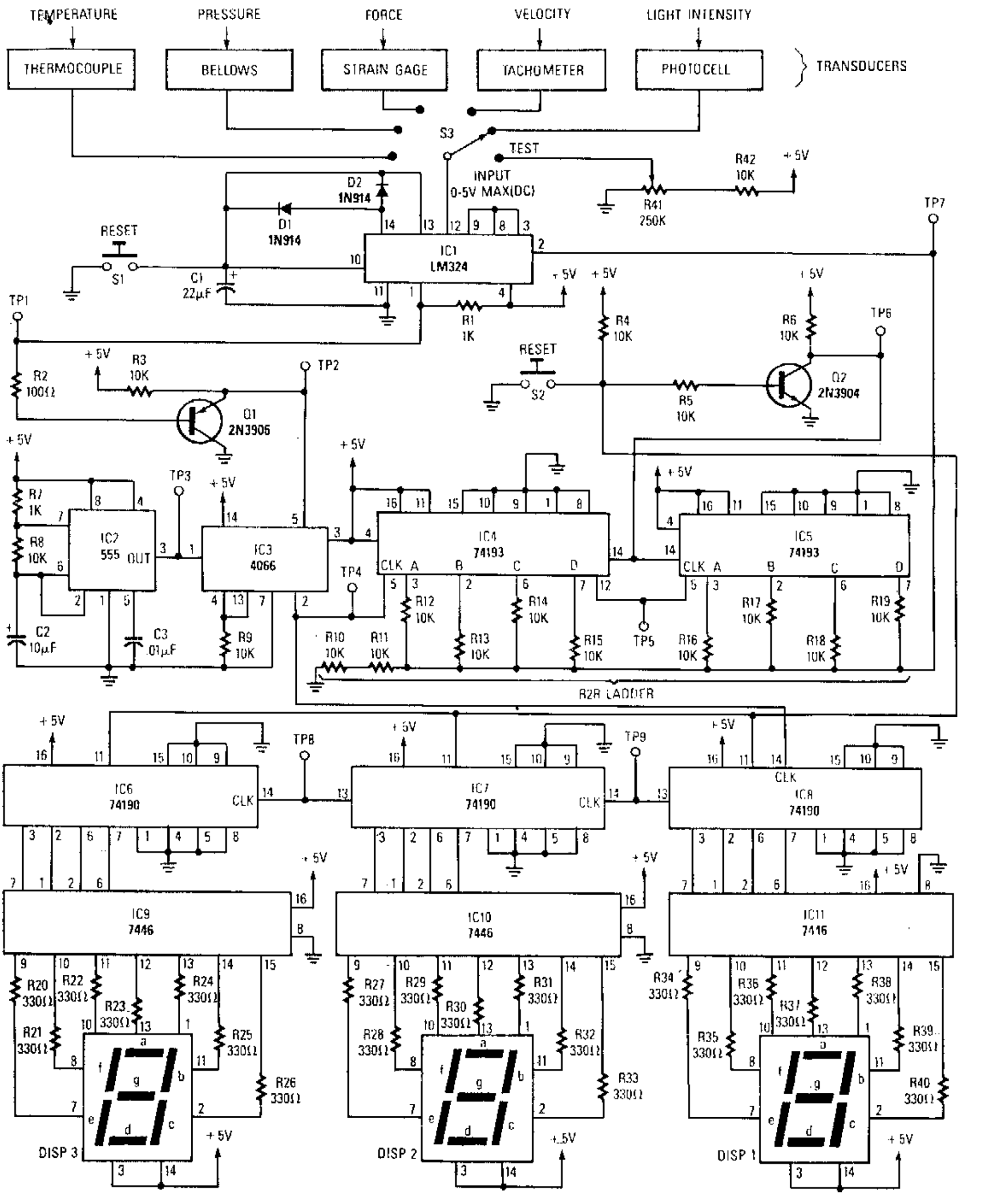
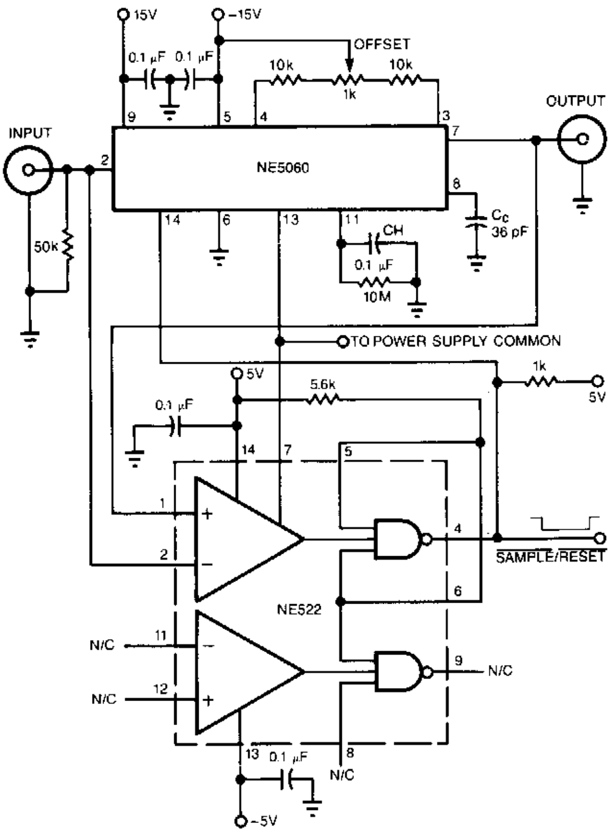


Fig. 25-14

DIGITAL PEAK DETECTOR CONT.

The peak detector tracks and holds, using the charge-storing ability of a capacitor, the highest output voltage from a transducer. Initially, the voltage on the inverting input of the comparator is at ground level. As a small voltage (0 – 5 V) is captured by the peak detector and presented to the comparator's noninverting input, the output will swing high, which asserts the bilateral switch; clock pulses now pass through the switch to clock both the BCD and binary counters. The outputs of the binary counters are connected to an R2R ladder network, which functions as a digital-to-analog converter. As the binary count increases, the R2R ladder voltage also increases until it reaches a point slightly above the voltage of the peak detector; at that instant, the comparator output swings low, which disables the bilateral switch and stops the counters. The number displayed on the 7-segment LED's will represent a value equivalent to the transducer's output.

HIGH-BANDWIDTH PEAK DETECTOR



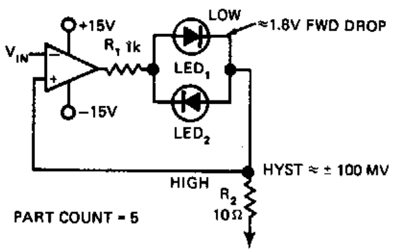
The high-speed peak detector uses a highly accurate, fast s/h amplifier controlled by a high-speed comparator. The s/h amplifier holds the peak voltage, until the comparator switches the amp to its sample mode, to capture a new, higher voltage level. The circuit handles all common-wave shapes and exhibits 5% accuracy from 50 Hz to 2 MHz.

The comparator's output decreases when the input signal exceeds the value of the currently held output. This transition puts the s/h amplifier into sample mode. Once the output reaches the value of the input, or the input signal falls below the output's level, the comparator's output increases; the high output brings the s/h amplifier back to the hold mode, thereby holding the peak value of the input signal. Reset the circuit by lowering the value of pin 4 of the NE522 comparator, which in turn allows the NE5060 s/h amplifier to acquire the input. The NE522 comparator has an open-collector output.

EDN

Fig. 25-15

NULL DETECTOR

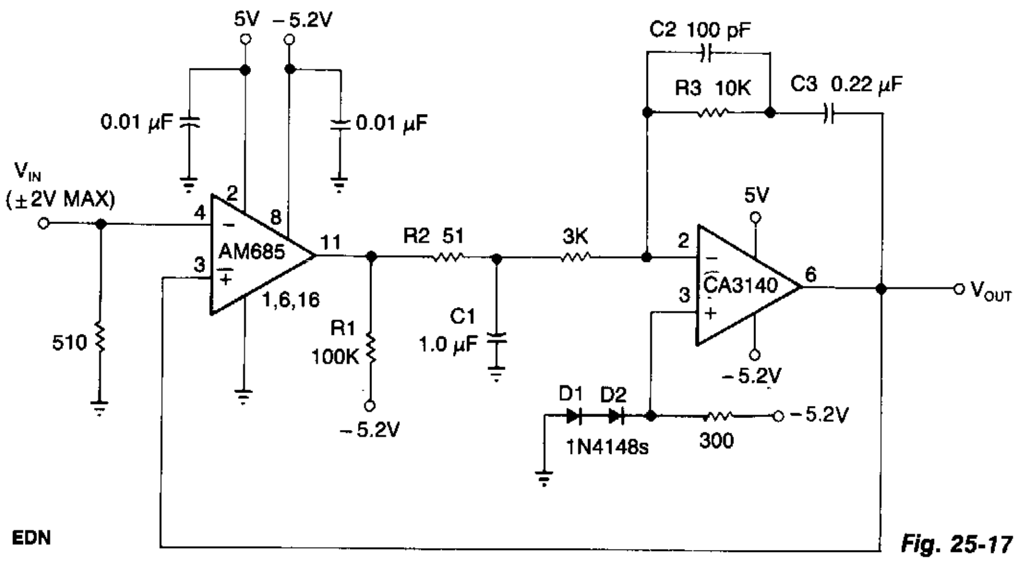


EDN

Fig. 25-16

In this indicating comparator circuit, R2 sets the hysteresis. If the 741 saturates at ± 12 V, the current in R1 will be approximately ± 10 mA if 0.1 V hysteresis is desired. Then 0.1 V/10 mA = 10Ω = *R*2.

WIDE-BANDWIDTH PEAK DETECTOR



EDN

Fig. 25-17

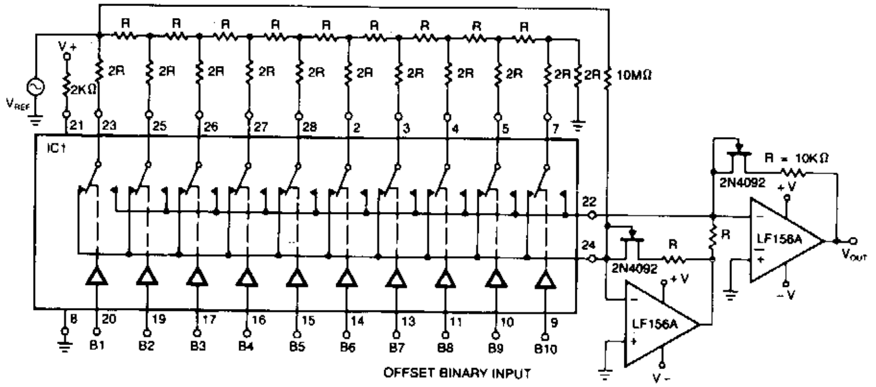
This circuit can detect the positive peaks for signal frequencies higher than 5 MHz. It yields ±1% accuracy for 400 mV to 4 V pk-pk signal amplitudes on sine, square, and triangular waveforms. The Am685 comparator output increases whenever V_{IN} is a greater negative voltage than V_{OUT}; the high comparator output, in turn, charges C1 in a positive direction. The CA3140 op amp amplifies the C1 voltage with respect to the ECL-switching-threshold voltage (−1.3 V) developed by diodes D1 and D2. For repetitive waveforms, each cycle boosts V_{OUT} until it equals the peak input value. The peak-detection process is aided by the comparator's open-emitter output, which allows C1 to charge rapidly through R2, but to discharge slowly through R2 and R1. Reducing the value of C1 shortens system-response times. Although the circuit can't detect negative-going peaks, it can be modified to measure the pk-pk value of bipolar signals that are symmetric about ground. To do so, divide V_{OUT} by 2 using two 1-KΩ resistors and feed the comparator $V_{OUT}/2$ rather than V_{OUT}.

26

Digital-to-Analog Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

10 Bit, 4 Quadrant Multiplying DAC
(Offset Binary Coding)



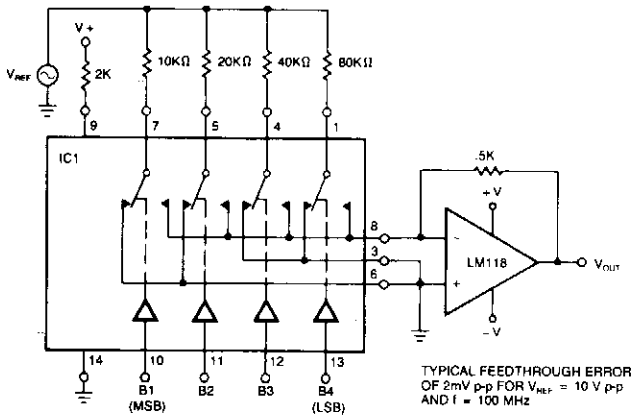
IC1: use five of either DG403, DG413, or DG423

Biopolar (Offset Binary)* Operation

DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1	$-V_{REF}(1-2^{-9})$
1 0 0 0 0 0 0 0 0 1	$-V_{REF}(2^{-9})$
1 0 0 0 0 0 0 0 0 0	0
0 1 1 1 1 1 1 1 1 1	$V_{REF}(2^{-9}$
0 0 0 0 0 0 0 0 0 1	$V_{REF}(1-2^{-9})$
0 0 0 0 0 0 0 0 0 0	V_{REF}

NOTE: 1 LSB = 2^{-9} V_{REF}
*Complementing B1 (MSB) will give 2's complement coding.

4 Bit Multiplying Current Switch D/A



IC1: use two of either DG403, DG413, or DG423

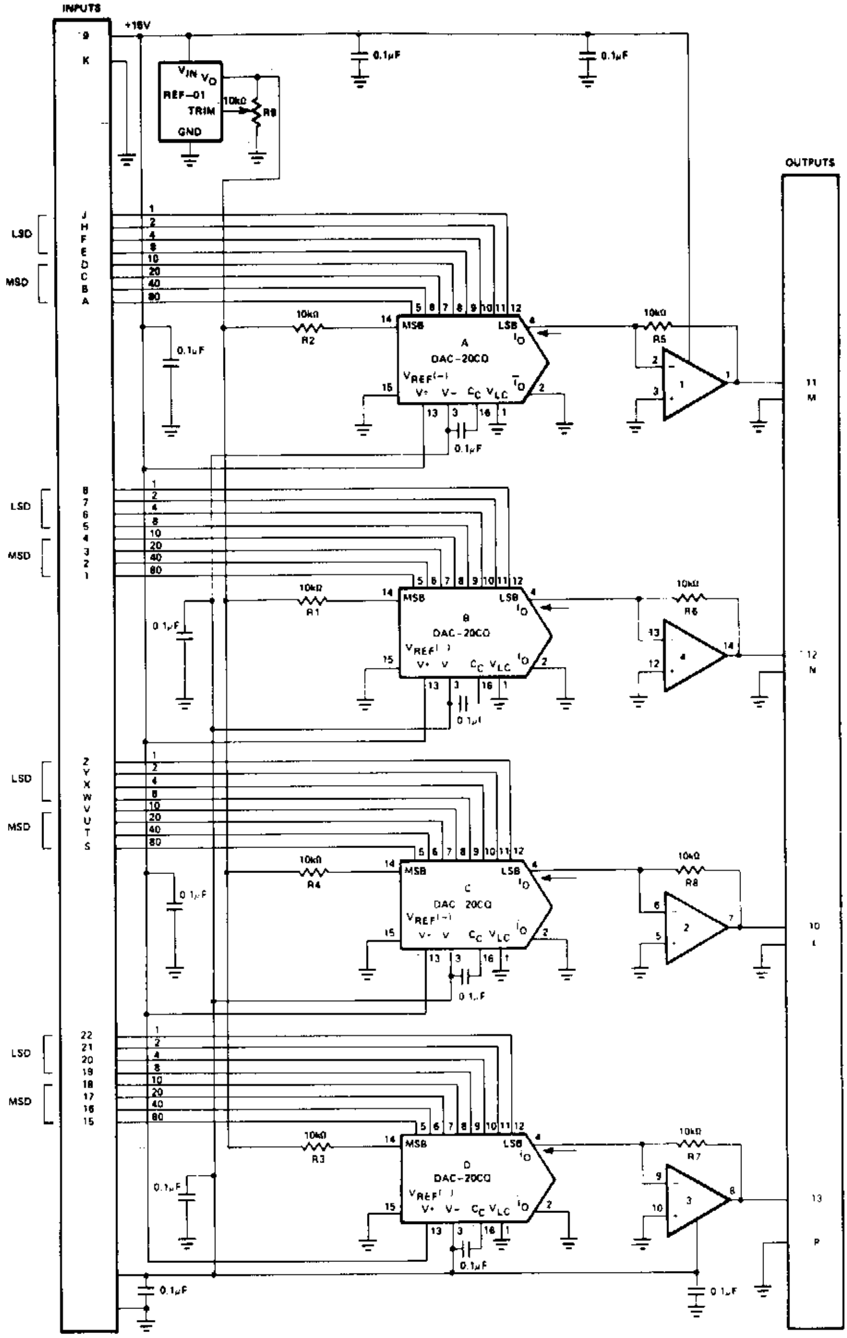
SILICONIX

Fig. 26-1

The following applications circuits are intended to illustrate the following points:

- A 2-KΩ resistor should be in series with V+ to limit supply current with negative ringing of the bit inputs
- Temperature compensation for $R_{DS(on)}$ can be provided in the feedback path of the op amp
- Bipolar reference voltages can be used in all configurations

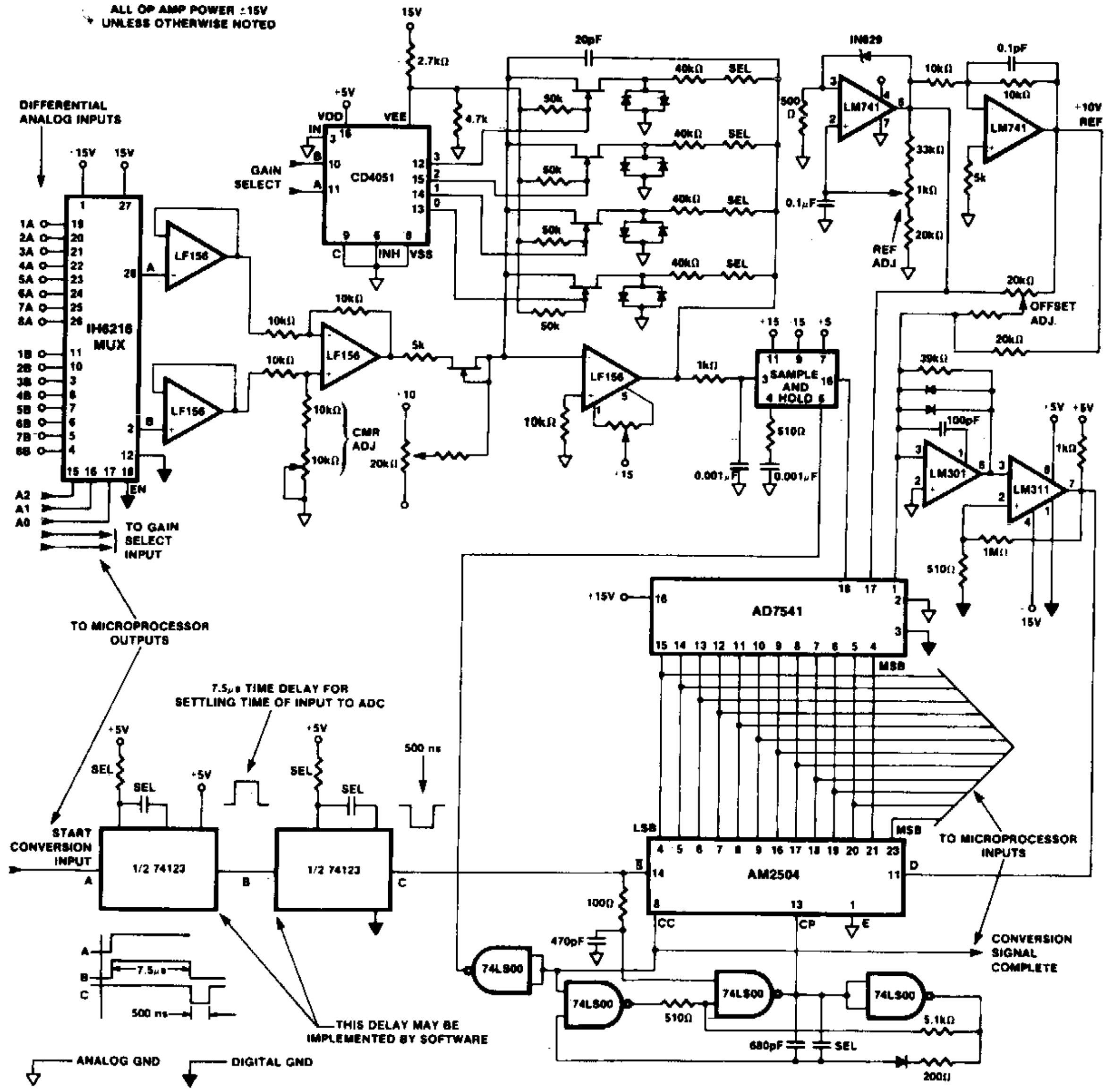
FOUR-CHANNEL D/A OUTPUT AMPLIFIER



PRECISION MONOLITHICS INC.

Fig. 26-2

12-BIT BINARY 2s COMPLEMENT D/A CONVERSION SYSTEM

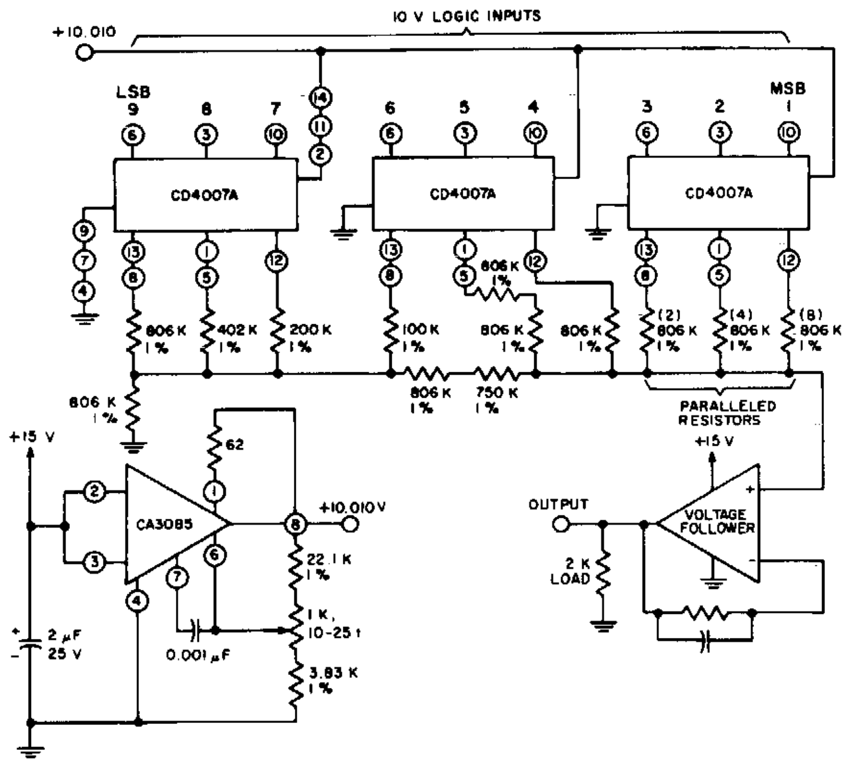


INTERSIL

Fig. 26-3

The front end of the DAC is configured differentially using dual eight-input IC multiplexer 1H6216 and three LM156 op amps. Following the differential amplifier is the programmable gain stage discussed earlier, with a low-pass filter on the output feeding the IH5110 sample and hold amplifier. The output of the IH5110 is connected to the comparator input, – input LM301, through the internal 10-KΩ feedback resistor of the 7541 multiplying d/a converter. The AD7541, along with a ±10-V reference and successive approximation logic, make up the 2's complement a/d converter.

9-BIT CMOS D/A CONVERTER



BIT	REQUIRED RATIO-MATCH
1	STANDARD
2	±0.1%
3	±0.2%
4	±0.4%
5	±0.8%
6–9	±1% ABS

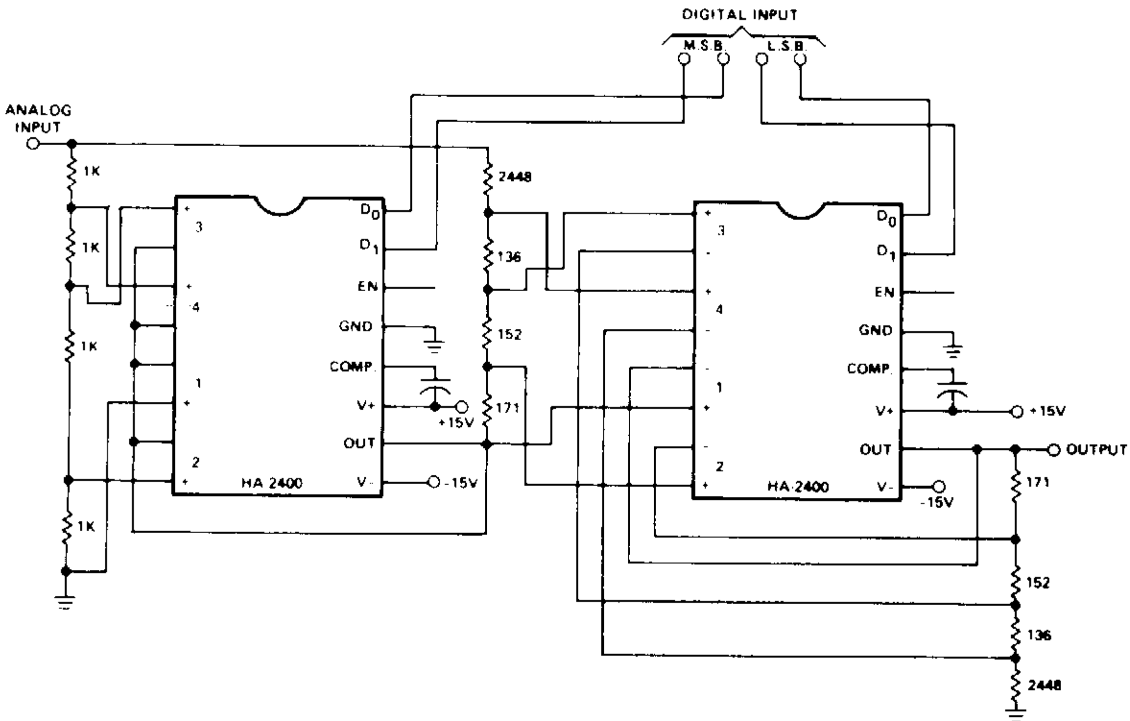
ALL RESISTANCES IN OHMS

INTERSIL

Fig. 26-4

Three CD4007A IC packages perform the switch function using a 10-V logic level. A single 15-V supply provides a positive bus for the follower amplifier and feeds the CA3085 voltage regulator. The scale adjust function is provided by the regulator output control, which is set to a nominal 10 V in this system. The line-voltage regulation (approximately 0.2%) permits 9-bit accuracy to be maintained with a variation of several volts in the supply. System power consumption ranges between 70 and 200 mW; a major portion is dissipated in the load resistor and op amp. The regulated supply provides a maximum current of 440 μA of which 370 μA flows through the scale adjusting. The resistor ladder is composed of 1% tolerance metal-oxide film resistors. The ratio match between resistance values is in the order of 2%. The follower amplifier has the offset adjustment nulled at approximately a 1 V output level.

MULTIPLYING D/A CONVERTER



HARRIS

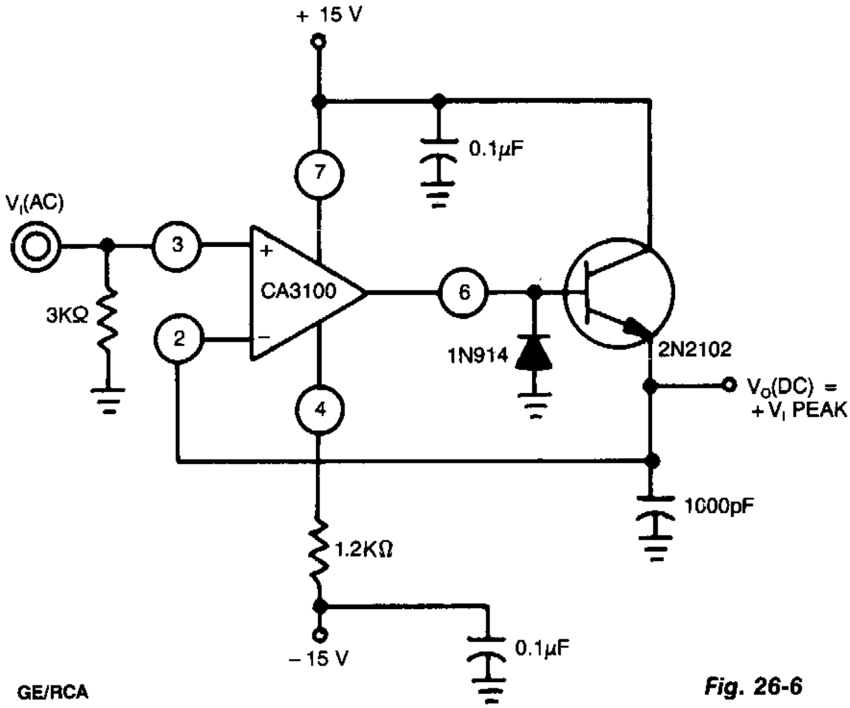
Fig. 26-5

The circuit performs the function:

$$V_{OUT} = V_{IN} \times \frac{N}{16}$$

where N is the binary number from 0 to 15 formed by the digital input. If the analog input is a fixed dc reference, the circuit is a conventional 4-bit D to ac signal, in which case the output is the product of the analog signal and the digital signal. The circuit on the left is a programmable attenuator with weights of 0, 1/4, 1/2, or 3/4. The circuit on the right is a noninverting adder, which adds weights to the first output of 0, 1/16, 1/8, or 3/16. If four quadrant multiplication is required, place a phase selector circuit in series with either the analog input or output. The D_O input of that stage becomes the + or − sign bit of the digital input.

POSITIVE PEAK DETECTOR



GE/RCA

Fig. 26-6

This peak detector uses a CA3100 BiMOS op amp as a wide-band noninverting amplifier to provide essentially constant gain for a wide range of input frequencies. The IN914 clips the negative half of V_{IN} $(R4)/(R3)$ $(R5)$. A 500-μA load current is constant for all load values and the output reflects only positive input peaks.

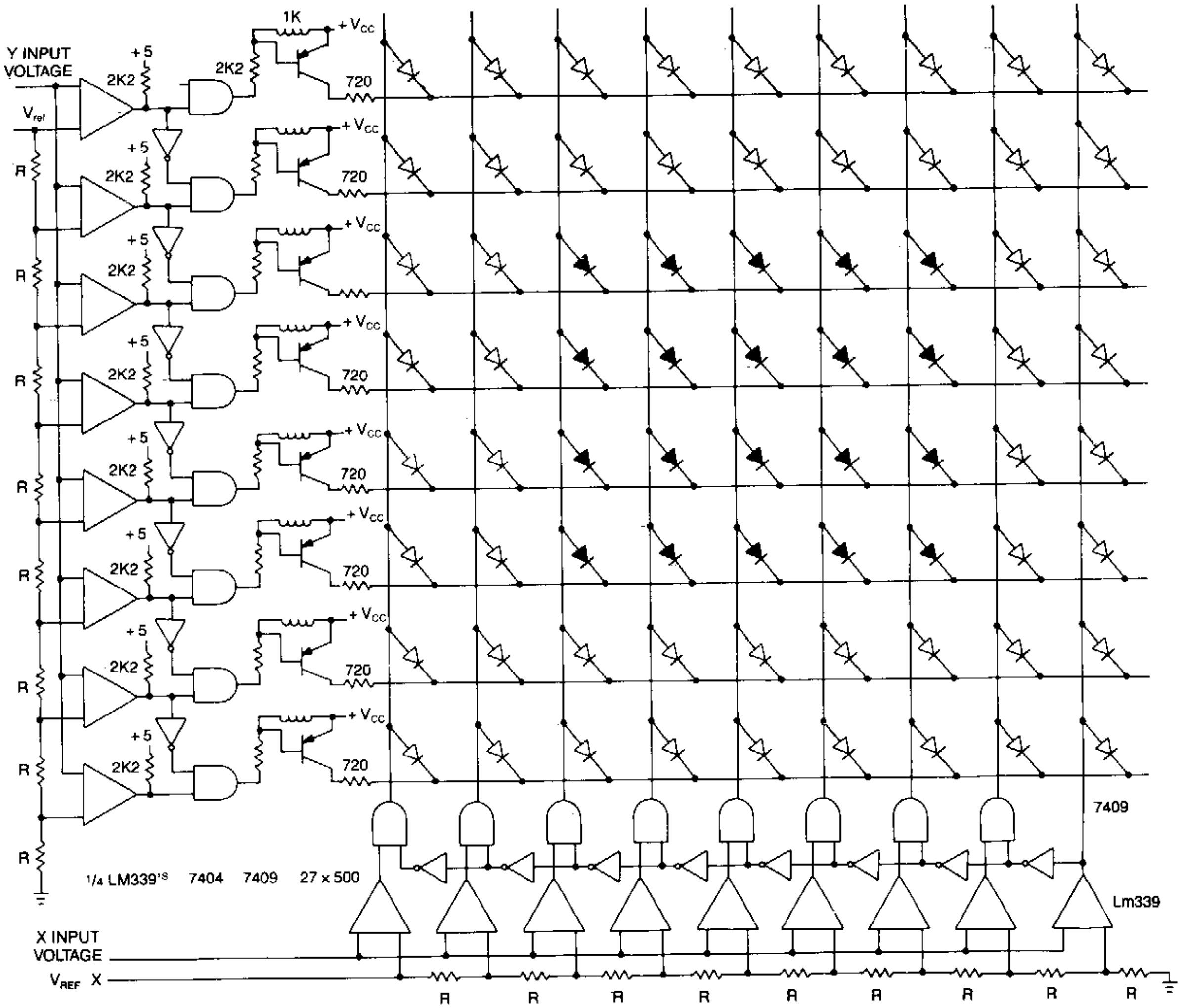
27

Display Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Two-Variable LED Matrix Display

TWO-VARIABLE LED MATRIX DISPLAY



ELECTRONIC ENGINEERING

Fig. 27-1

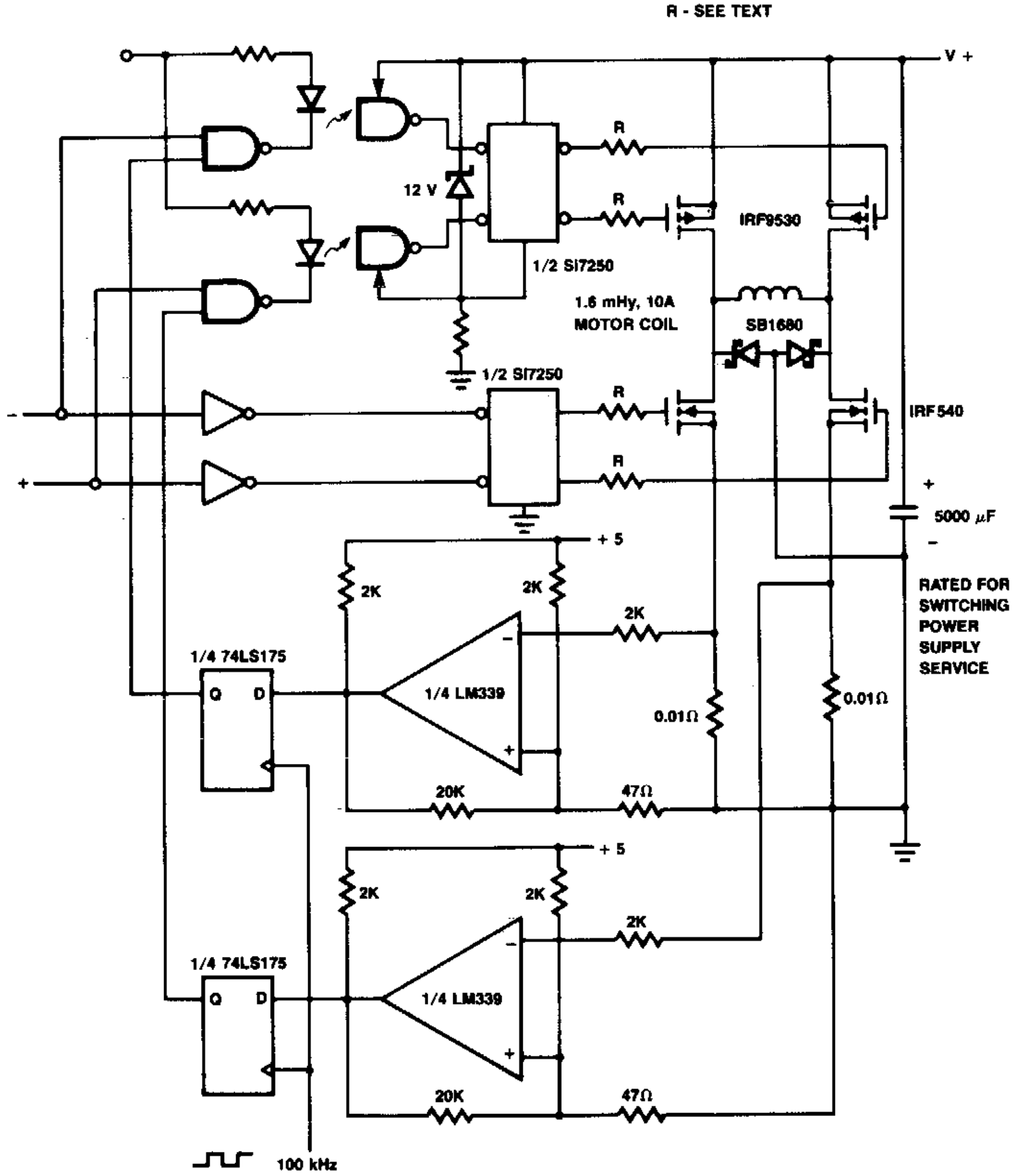
This matrix can show the values of two variables, for example, frequency and voltage. The display is a graph made from a matrix of LEDs. The LEDs on each axis are color coded, red for *out of tolerance* and green for *in*, forming a red band around the inner green rectangle. The two input voltages proportional to the functions being measured are presented to the two columns of comparators. The other comparator input is a reference voltage derived from resistor ladder R1 to R_X. The output of each row of comparators is processed with an inverted and an AND gate to allow only one active output for any input value. The LED at the intersection of the active drives shows the relationship of the two inputs. The advantage of this display is the ease in reading, modification, and also its small size. All comparators are LM339 quads.

28

Drive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PRACTICAL CURRENT-LIMITING COIL DRIVER



SILICONIX

Fig. 28-1

The p-channel devices are switched off by current sensors when the coil current reaches 10 A. The operation is similar to that of a switching-type power supply. The Schottky diodes and resistors are for spike protection.

LINE-SYNCHRONIZED DRIVER

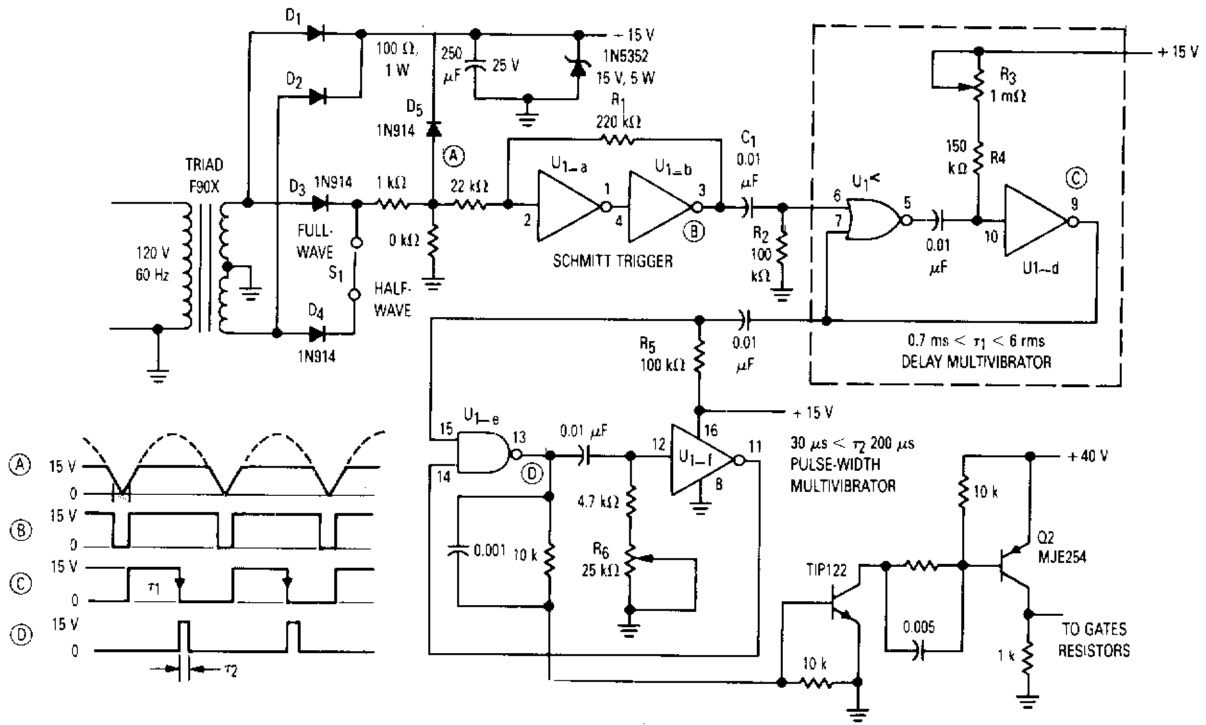
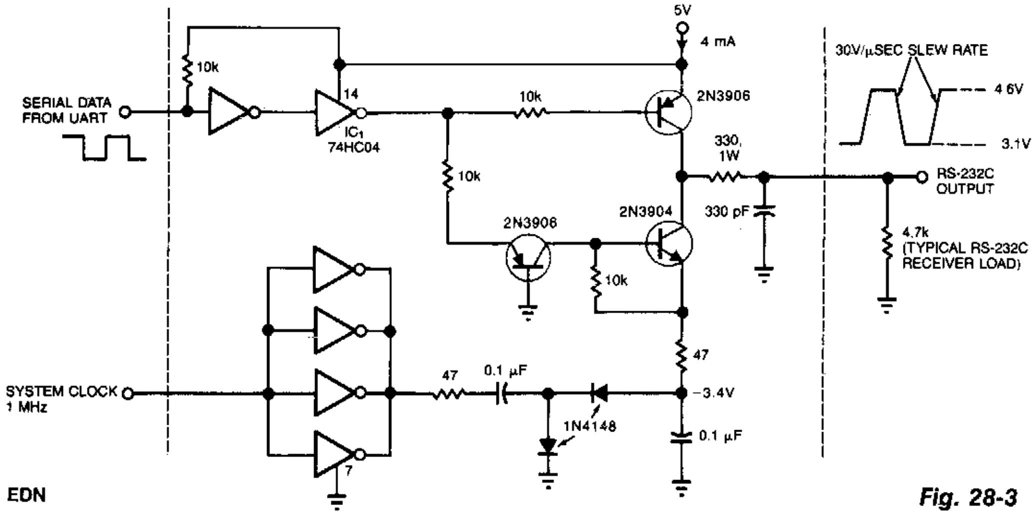


Fig. 28-2

The gate drive that phase controls the four parallel SCRs is accomplished with complementary MOS hex gate MC14572 and two bipolar transistors. This adjustable line-synchronized driver permits SCR conduction from near zero to 180 degrees. A Schmitt trigger clocks a delay monostable multivibrator that is followed by a pulse-width monostable multivibrator. Line synchronization is achieved through the half-wave section of the secondary winding of the full-wave, center-tapped transformer. This winding also supplies power to the circuit through rectifiers D1 and D2.

LOW-POWER RS-232C DRIVER

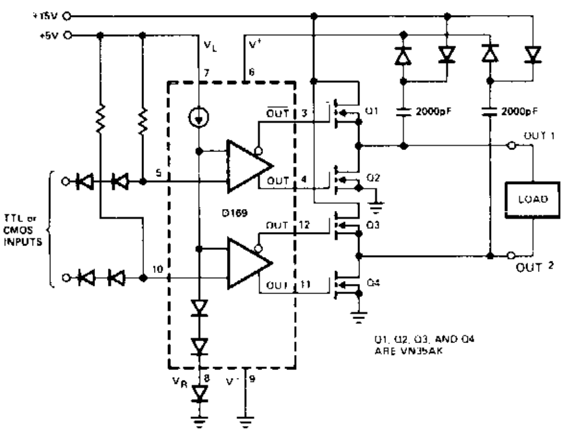


EDN

Fig. 28-3

This circuit draws only 4 mA from a 5-V supply while driving a standard RS-232C receiver. The system clock drives a dc-dc converter that produces −3.4 V. The frequency can range from 0.5 to 8 MHz, but a range of 0.5 to 1 MHz will minimize power dissipation. The circuit output withstands direct shorts to ground or to either of the supplies (±12 V max). In place of the 74HC04 high-speed CMOS driver shown, you may want to substitute miscellaneous spare gates; one noninverting buffer, for example, can replace the two inverting types that receive the UART signal.

TOTEM-POLE DRIVER WITH BOOTSTRAPPING



SILICONIX

Fig. 28-4

When driving MOSPOWER in a totem-pole output configuration, it is necessary to have the gate voltage 10 to 15 V positive with respect to the source in order to handle load currents near the MOSPOWER maximum ratings. The D169 lends itself to bootstrapping because of its high-voltage ratings. In the circuit shown, the voltage on the 2000-pF bootstrap capacitors is applied via diode OR gates to the V+ terminal. Therefore, regardless of which output is high, 30 V is present at V+. Maximum switching frequency is determined by the input capacitance of the MOSPOWER transistors used.

29

Fiber Optics Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

FIBER OPTIC TRANSMITTER

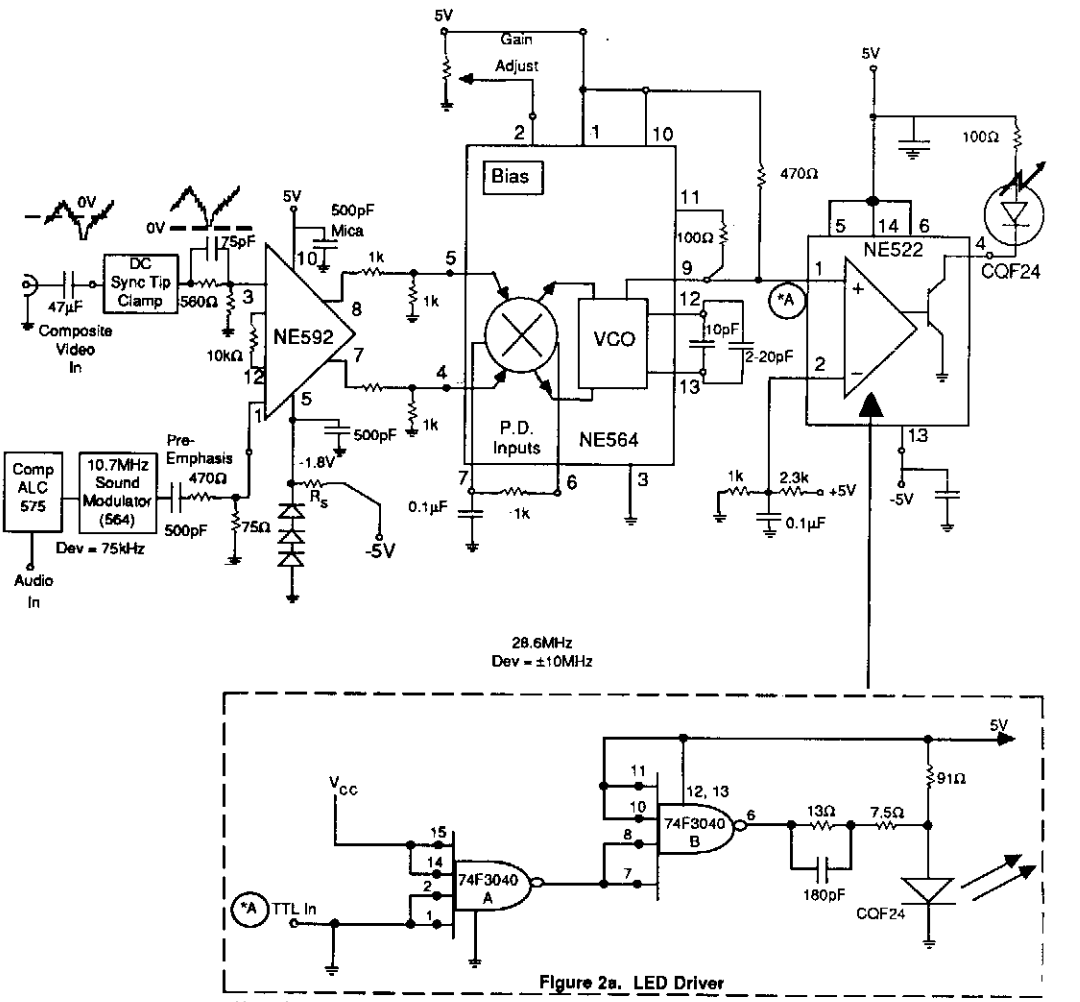


Figure 2a. LED Driver

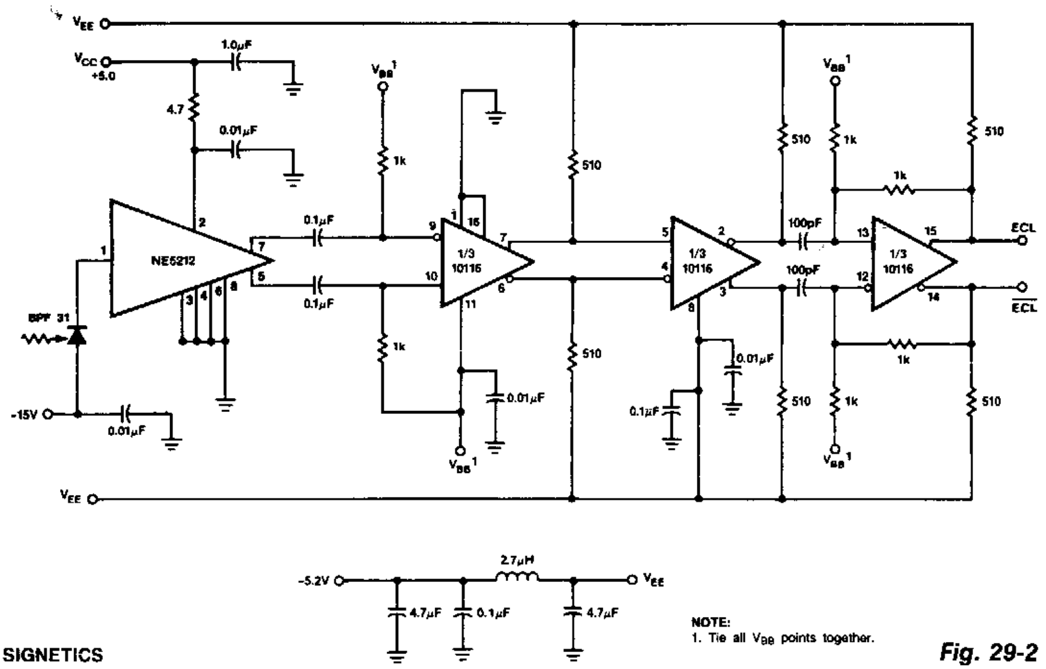
* Note: An alternate LED driver which uses the 74F3040 line driver was incorporated in this particular application example. The 74F3040 has a higher current rating, but not the variable threshold capabilities of the NE522. The LED diode is operated in the saturated on-off mode for best signal to noise.

SIGNETICS

Fig. 29-1

This receiver circuit consists of wideband differential amplifier NE592, VCO NE564 and LED driver NE522—the high-speed comparator. The video signal is ac coupled into the modulator preamplifier and followed by a sync tip clamp to provide dc restoration on the composite video signal and to prevent variation of modulation deviation with varying picture content. A video signal level of 250 to 300 mV peak is required to maintain optimum picture modulation. Frequency compensation (preemphasis) is inserted in the form of a passive rc lead network at the input to the NE592 differential amplifier. The main FM modulator consists of an NE564 used only as a linear wideband VCO, but the other sections of the device are not used. Differential dc coupling to the VCO terminals is attained via the loop filter terminals, pins 4 and 5.

DIGITAL FIBER OPTIC RECEIVER

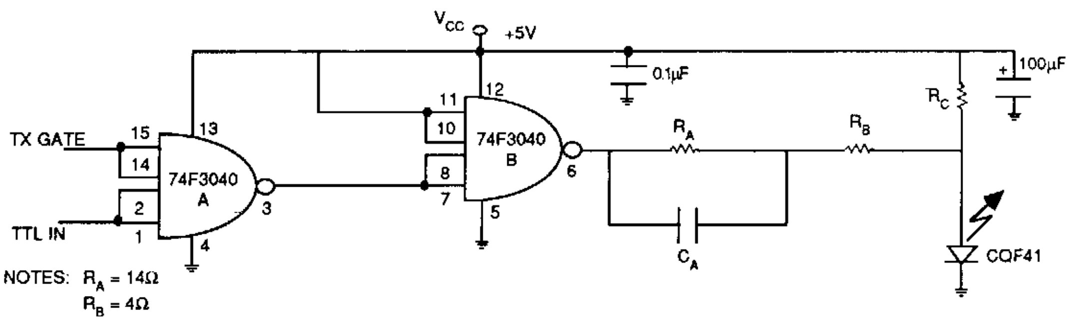


SIGNETICS

Fig. 29-2

This receiver uses the NE5212, the Signetics 10116 ECL line receiver, and the Phillips/Amperex BPF31 pin diode. The circuit is a capacitor-coupled receiver and utilizes positive feedback in the last stage to provide the hysteresis. The amount of hysteresis can be tailored to the individual application by changing the values of the feedback resistors to maintain the desired balance between noise immunity and sensitivity. At room temperature, the circuit operates at 50-M baud with a BER of 10E-10 and over the automotive temperature range at 40-M baud with a BER of 10E-9. Higher speed experimental diodes have been used to operate this circuit at 220-M baud with a BER of 10E-10.

50-Mb/s FIBER OPTIC LED DRIVER



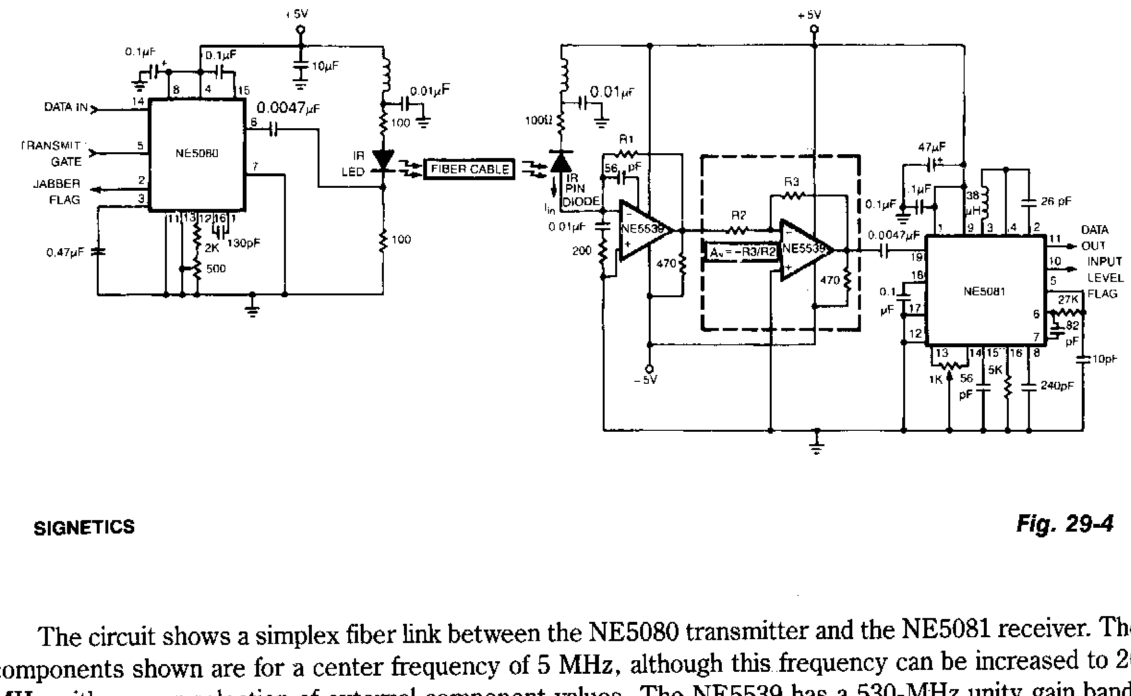
NOTES: $R_A = 14\Omega$
$R_B = 4\Omega$
$R_C = 90\Omega$
$C_A = 250pF$

SIGNETICS

Fig. 29-3

The pull-up transistor of the totem-pole output is used to turn on the LED and the pull-down transistor is used to turn off the LED. The lower impedance and higher current handling capability of the saturated pull-down transistor is used as an effective method of transferring the charge from the LED's anode to ground as its dynamic resistance increases during turn-off. The slightly higher output impedance of the pull-up stage ensures that the LED is not over peaked during the less difficult turn-on transition. This asymmetric current handling capability of the output stage with its variable impedance substantially reduces the pulse-width distortion and long-tailed response. As the signal propagates through two NAND gates, each transition passes through the high-to-low and low-to-high transition once, normalizing the total propagation delay through the circuit.

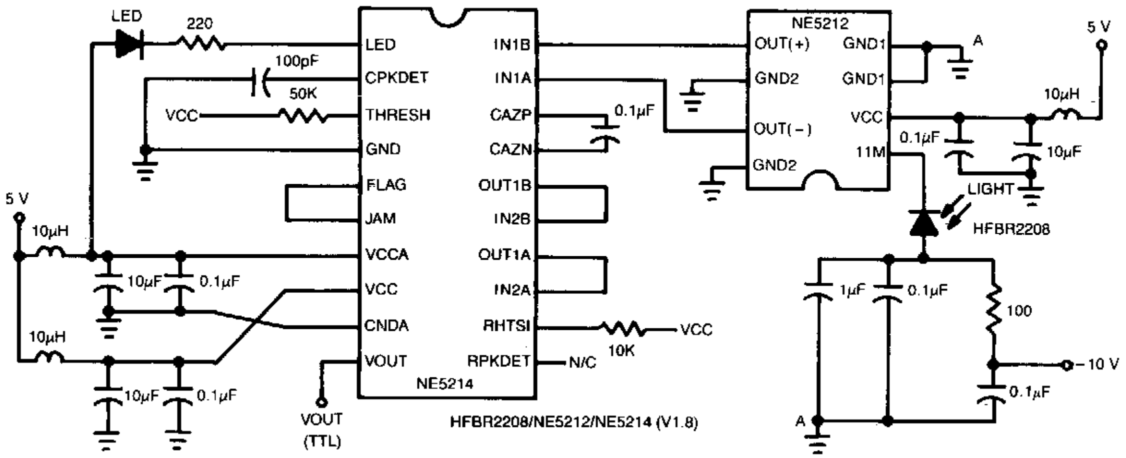
FIBER OPTIC LINK



SIGNETICS *Fig. 29-4*

The circuit shows a simplex fiber link between the NE5080 transmitter and the NE5081 receiver. The components shown are for a center frequency of 5 MHz, although this frequency can be increased to 20 MHz with proper selection of external component values. The NE5539 has a 530-MHz unity gain bandwidth which could limit maximum operating frequencies in some systems. Since the NE5081 can adequately accept signals below 10 mV at 5-MHz carrier, the gain stage within the dashed lines can be eliminated if the attenuation in the link is low. If the gain stage is used, be mindful of the bandwidth tradeoff at higher gains. Refer to the NE5539 data sheet for details.

LOW-COST 100-M BAUD FIBER OPTIC RECEIVER



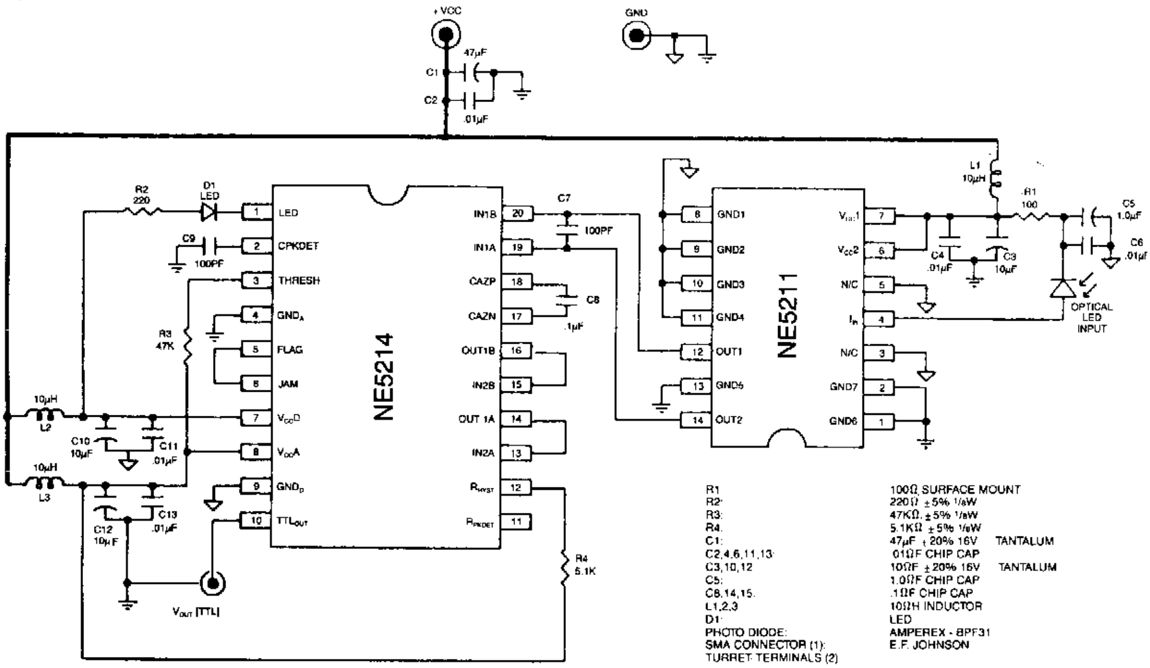
SIGNETICS

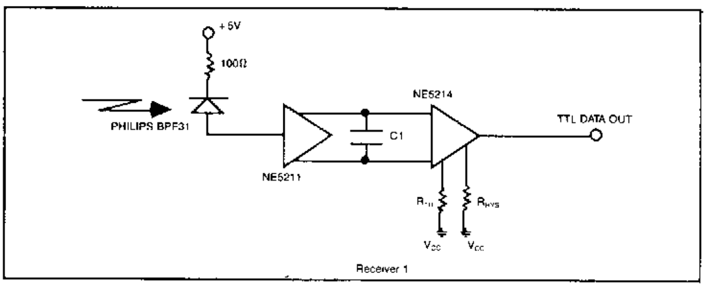
Fig. 29-5

This two-chip receiver with minimum external component count has been designed for low-cost fiber optic applications to 100-M baud (50 MHz). The receiver is divided into pre- and postamplifier ICs for increased stability. The preamplifier IC features low noise with a differential transresistance design. The postamplifier IC incorporates an auto-zeroed first stage with noise shaping, high-gain symmetrical-limiting amplifier, and a matched rise/fall time TTL output buffer. A wide-band full-wave rectifier functions as a link-status indicator. To ensure stability, a surface mount, small outline (SO), package is used. The received signal in the −35 dBm optical (average) to −9 dBm range is converted into a small unipolar current by the pin diode. The pin diode then feeds its signal current to a preamplifier, such as the NE5212. The preamplifier output is fed to a high-gain limiting amplifier, simply known as the post amp.

The NE5214/NE5217 postamplifiers are low-cost ICs that provide up to 60 dB of gain at 50 MHz to bring mV level signals up to TTL levels. The postamplifier IC incorporates an auto-zeroed first stage with noise shaping, a high-gain symmetrical-limiting amplifier, and a matched rise/fall time TTL output buffer. A secondary amplifier chain functions as a link-status indicator.

50-Mb/s FIBER OPTIC RECEIVER





SIGNETICS

Fig. 29-6

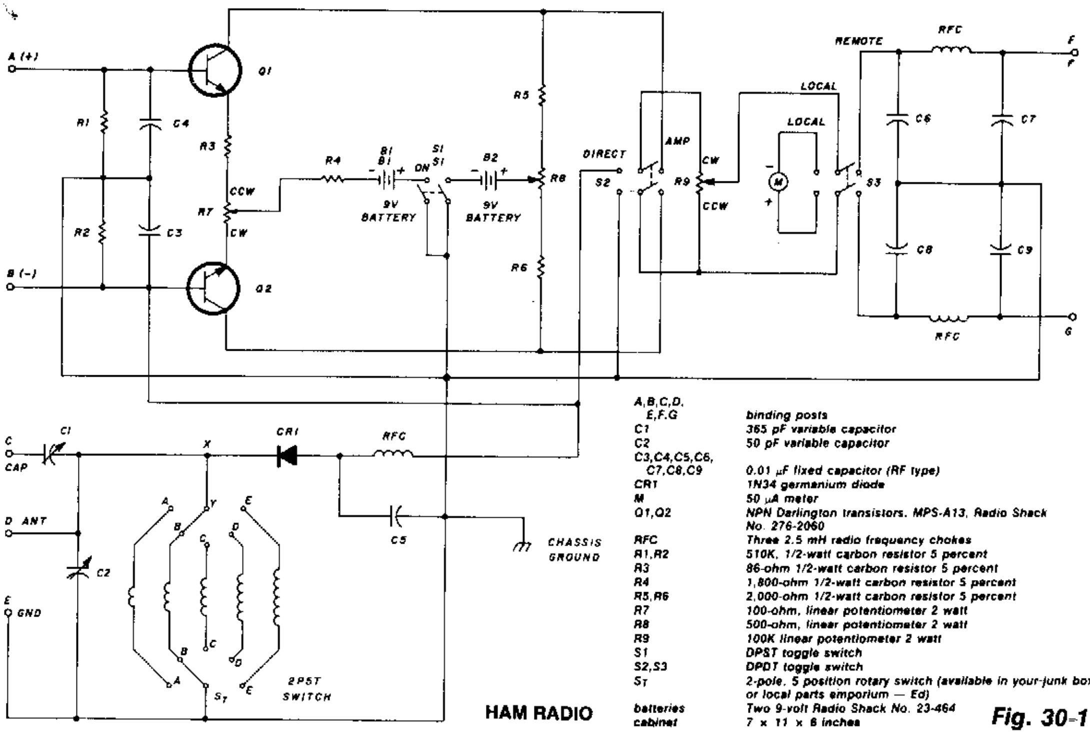
The optical signal is coupled to the pin diode. Current flowing in the diode also flows into the input of the NE5211 preamplifier. The preamplifier is a fixed-gain block that has a 28-KΩ differential transimpedance and does a single-ended to differential conversion. With the signal in differential form, greater noise immunity is assured. The second stage, or postamplifier NE5214, includes a gain block, auto-zero detection, and limiting. The auto-zero circuit allows dc coupling of the preamplifier and the postamplifier and cancels the signal dependent offset because of the optical-to-electrical conversion. The auto-zero capacitor must be 1000 pF or greater for proper operation. The peak detector has an external threshold adjustment, R_{TH}, allowing the system designer to tailor the threshold to the individual's need. Hysteresis included to minimize jitter introduced by the peak detector, and an external resistor, R_{HYS}, is used to set the amount of hysteresis desired. The output stage provides a single-ended TTL data signal with matched rise and fall times to minimize duty-cycle distortion.

30

Field-Strength Meters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SENSITIVE FIELD-STRENGTH METER



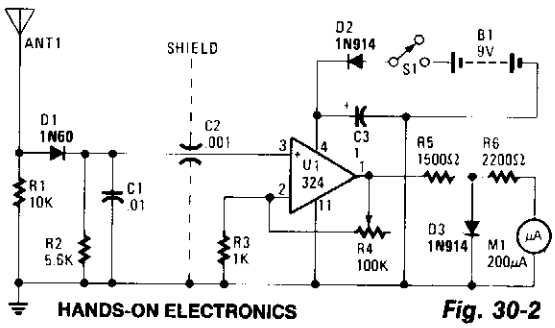
A,B,C,D, E,F,G	binding posts
C1	365 pF variable capacitor
C2	50 pF variable capacitor
C3,C4,C5,C6, C7,C8,C9	0.01 μF fixed capacitor (RF type)
CR1	1N34 germanium diode
M	50 μA meter
Q1,Q2	NPN Darlington transistors. MPS-A13, Radio Shack No. 276-2060
RFC	Three 2.5 mH radio frequency chokes
R1,R2	510K, 1/2-watt carbon resistor 5 percent
R3	86-ohm 1/2-watt carbon resistor 5 percent
R4	1,800-ohm 1/2-watt carbon resistor 5 percent
R5,R6	2,000-ohm 1/2-watt carbon resistor 5 percent
R7	100-ohm, linear potentiometer 2 watt
R8	500-ohm, linear potentiometer 2 watt
R9	100K linear potentiometer 2 watt
S1	DPST toggle switch
S2,S3	DPDT toggle switch
S_T	2-pole, 5 position rotary switch (available in your-junk box or local parts emporium — Ed)
batteries	Two 9-volt Radio Shack No. 23-464
cabinet	7 × 11 × 6 inches

HAM RADIO

Fig. 30-1

The two-pole, five-position switch, coils and 365-pF variable capacitor cover a range from 1.5 to 30 MHz. The amplifier uses Darlington npn transistors whose high beta, 5000, provides high sensitivity with S1 used as the amplifier on/off switch. Switch S2 in the left position allows the output of the 1N34 diode to be fed directly into the 50-μA meter (M) for direct reading. When S2 is in the right position, the amplifier is switched into the circuit. Switch S3 is for local or remote monitoring. At full gain setting, the input signal is adjusted to give a full-scale reading of 50 mA on the meter. Then with the amplifier switched out of the circuit, the meter reading drops down to about 0.5 mA. A 2.5-mH rf choke and capacitors C3, C4, and C5 effectively keep rf out of the amplifier circuit.

FIELD-STRENGTH METER



HANDS-ON ELECTRONICS

Fig. 30-2

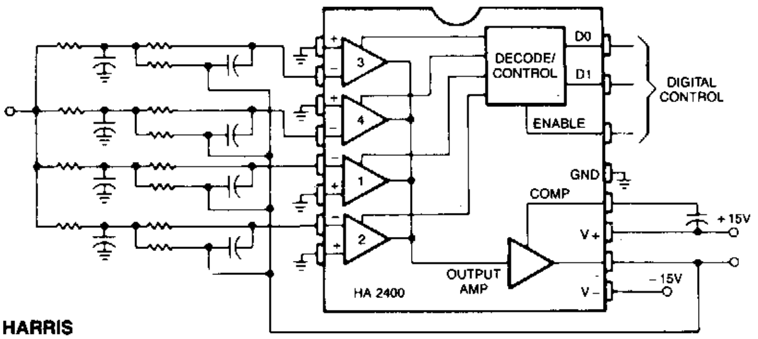
The untuned, but amplified FSM can almost sense that mythical flea's whisper—from 3 through 148 MHz no less—and yet, is so immune to overload that the meter pointer won't pin. The key to the circuit is the amplifier, a 324 quad op amp, of which only one section is used. It's designed for a single-ended power supply, will provide at least 20-dB dc gain, and the output current is self-limiting. The pointer can't be pinned.

31

Filter Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PROGRAMMABLE ACTIVE FILTERS

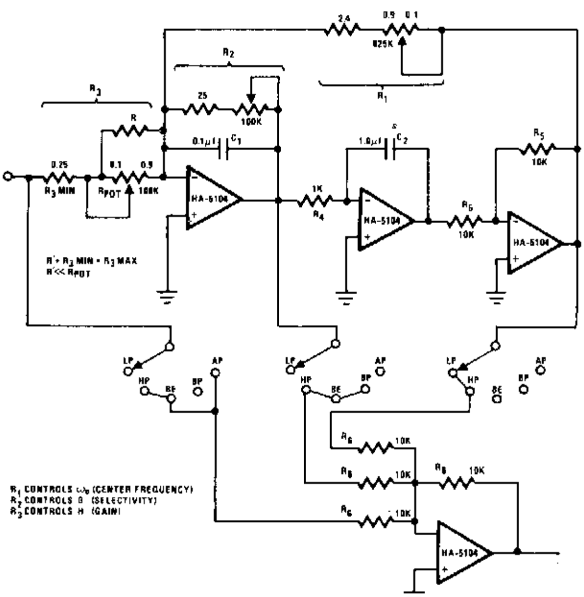


HARRIS **Fig. 31-1**

This is a second-order, low-pass filter with programmable cutoff frequency. This circuit should be driven from a low-source impedance since there are paths from the output to the input through the unselected networks. Virtually any filter function which can be constructed with a conventional op amp can be made programmable with the HA-2400.

A useful variation would be to wire one channel as a unity gain amplifier, so that one could select the unfiltered signal, or the same signal filtered in various manners. These could be cascaded to provide a wide variety of programmable filter functions.

BIQUAD AUDIO FILTER



The biquad offers a universal filter with ω_0, Q, and gain "orthogonally" tuned.

HARRIS **Fig. 31-2**

This universal filter offers low-pass, high-pass, bandpass, band elimination, and all-pass functions. The Biquad consists of two successive integration stages followed by an inverting stage. The entire group has a feedback loop from the front to the back consisting of R1 which is chiefly responsible for controlling the center frequency, ω_o. The first stage of integration is a *poor* integrator because R2 limits the range of integration. R2 and C form the time constant of the first stage integrator with R3 influencing gain H almost directly. The band-pass function is taken after the first stage with the low-pass function taken after the third stage. The remaining filter operations are generated by various combinations of three stages.

The Biquad is orthogonally tuned, meaning that ω_o, Q, and gain H can all be independently adjusted. The component values known will allow ω_o to range from 40 Hz to 20 kHz. The other component values give an adequate range of operation to allow for virtually universal filtering in the audio region. ω_o, Q, and gain H can all be independently adjusted by tuning R1 through R3 in succession.

LOW-POWER ACTIVE FILTER WITH DIGITALLY SELECTABLE CENTER FREQUENCY

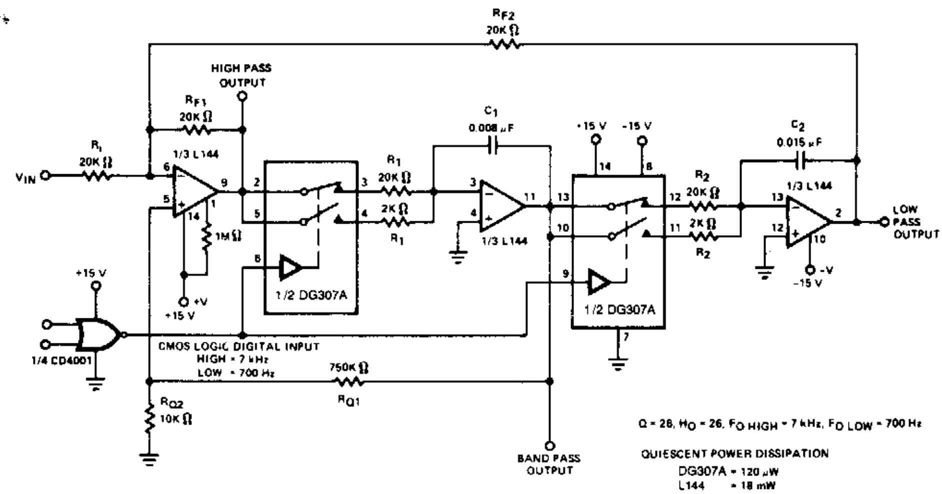


Table 1
Design Procedure for the State Variable Active Filter
Given: f_0 (Resonant Frequency),
H_0 (Gain at the Resonant Frequency) and Q_0

STANDARD DESIGN
(Assumes Infinte Op-Amp Gain)

1. CHOOSE $C_1 = C_2 = C$, A CONVENIENT VALUE
2. LET $R_1 = R_2 = R$
3. THEN $R = \dfrac{1}{2\pi \times f_o \times C}$
4. CHOOSE $R_{11} = R_{12} = KR$, WHERE R_{11}, R_{12} = A CONVENIENT VALUE

AND $K = \dfrac{H_o}{Q_o}$

IF H_o IS UNIMPORTANT (i.e., GAIN CAN BE ADDED BEFORE AND/OR AFTER THE FILTER), CHOOSE K = 1

5. LET R_{Q1} = A CONVENIENT VALUE
6. THEN $R_{Q2} = \dfrac{R_{Q1}}{(2 + K) \times Q_o - 1}$

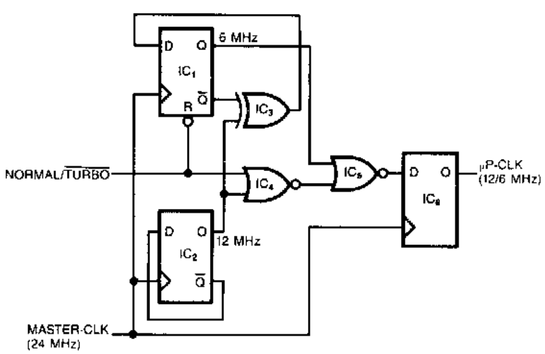
A (f_o) = THE NOMINAL OP AMP GAIN AT THE RESONANT FREQUENCY.

GBWP = THE NOMINAL GAIN-BANDWIDTH PRODUCT OF THE OPERATIONAL AMPLIFIER

SILICONIX *Fig. 31-3*

The switchable center frequency active filter allows a decade change in center frequency.

GLITCH-FREE TURBO CIRCUIT

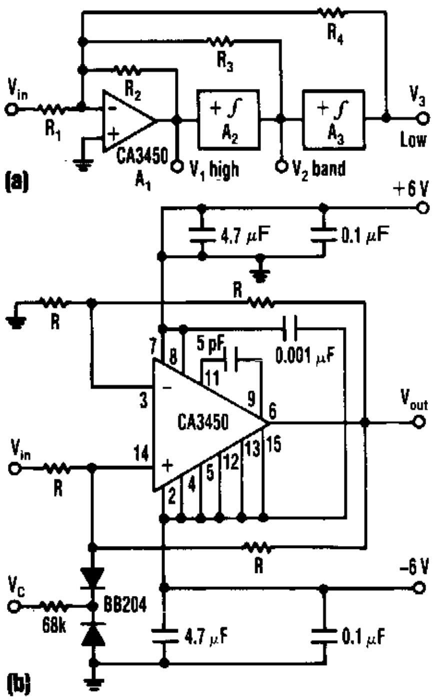


EDN *Fig. 31-4*

This simple circuit generates a dual-speed clock for personal computers. The circuit synchronizes your asynchronous switch inputs with the master clock to provide glitch-free transitions from one clock speed to the other. The dual-speed clock allows some programs to run at the higher clock speed in order to execute more quickly. Other programs—for example, programs that use loops for timing—can still run at the lower speed as necessary. The circuit will work with any master-clock frequency that meets the flip-flops' minimum-pulse-width specs.

The two D two flip-flops, IC1 and IC2, and an XOR gate, IC3, form a binary divider that develops the 6- and 12-MHz clocks. When the NT signal is low, the reset pin forces the 6-MHz output low. On the other hand, when the NT signal is high, IC3 blocks the 12-MHz output. Therefore, only one of the two clock signals passes through IC3 and gets clocked into IC6. Because the master-clk signal clocks IC6, asynchronous switching of the NT signal can't generate an output pulse shorter than 41 μs (1/24 MHz). Also, the synchronization eliminates glitches.

VOLTAGE-CONTROLLED FILTER



ELECTRONIC DESIGN *Fig. 31-5*

The control voltage V_C easily sets the cutoff frequency ω_o of this state-variable filter to any desired value, from about 1.7 MHz up to 5 MHz, with a BB 204 varicap and $R = 100\ K\Omega$. V_C can range from 0 to 28 V. This range changes the capacitance of the varicap from about 4 to 12 pF.

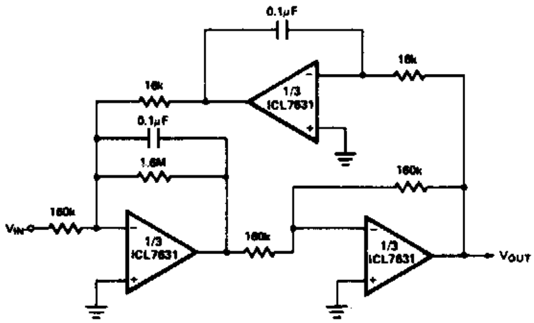
The circuit consists of input summing circuit A1 and two noninverting integrators, A2 and A3. Both the integrators and the summing-amplifier circuits use CA3450 op amps. With them, cutoff frequencies up to 200 MHz are possible.

The circuit's cutoff frequency, it's Q-factor, and gain G are simply:

$$\omega_o = 2/CR,\ Q = R_3/R_4,\ \text{and}\ G = R_4/R_1$$

For a given value for R4, say 10 KΩ, Q depends only upon the resistance of R3. The Q can be any value, even 100, independently of both ω_o and G. Similarly, the gain then depends only on the resistance of R1 and can also be set as high as 100.

SECOND-ORDER BIQUAD BANDPASS FILTER

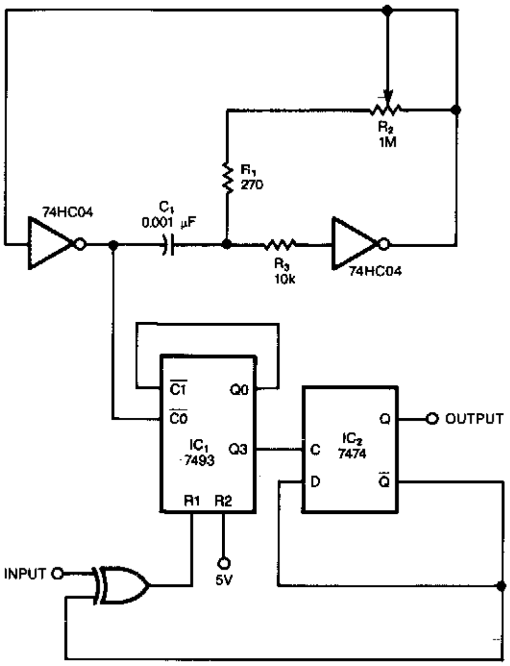


INTERSIL

Fig. 31-6

Note that I_Q on each amplifier might be different. $A_{VCL} = 10$, $Q = 100$, $f_o = 100$ Hz.

NOISY SIGNALS FILTER

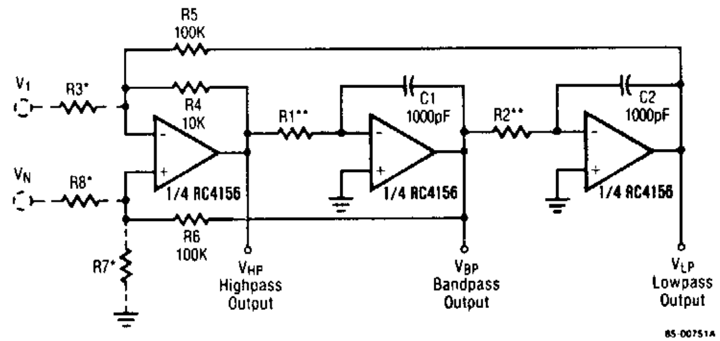


EDN

Fig. 31-7

This circuit filters noise, such as glitches and contact bounce, from digital signals. You can easily adjust the circuit for a wide range of noise frequencies. The circuit's output changes state only if the input differs from the output long enough for the counter to count eight cycles. If the input changes before the counter reaches its maximum count, the counter resets without clocking the output of flip-flop, IC2. You use R2 to set the frequency of the two-inverter CMOS oscillator, which clocks the counter. Simply adjust the oscillator such that its period is one-eighth that of the noise you want to eliminate.

STATE-VARIABLE ACTIVE FILTER



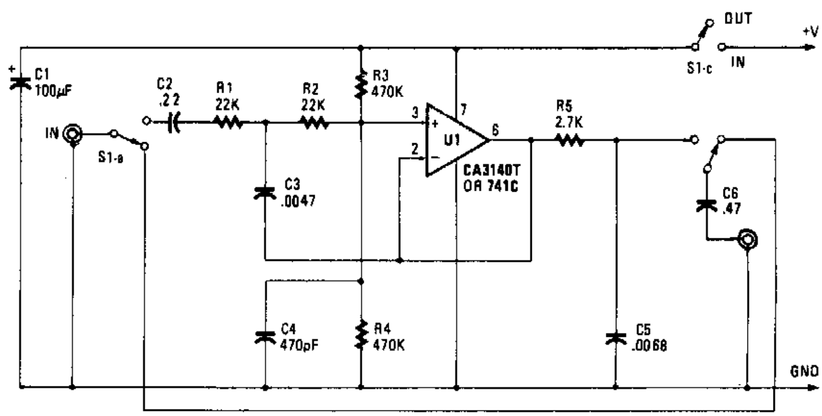
*Input connections are chosen for inverting or non-inverting response. Values of R3, R7, R8 determine gain and Q.
**Values of R1 and R2 determine natural frequency.

Reprinted with permission from Raytheon Co., Semiconductor Div. *Fig. 31-8*

A generalized circuit diagram of the two-pole state-variable active filter is shown. The state-variable filter can be inverting or noninverting and can simultaneously provide three outputs: low-pass, bandpass, and high-pass. A notch filter can be realized by adding one summing op amp.

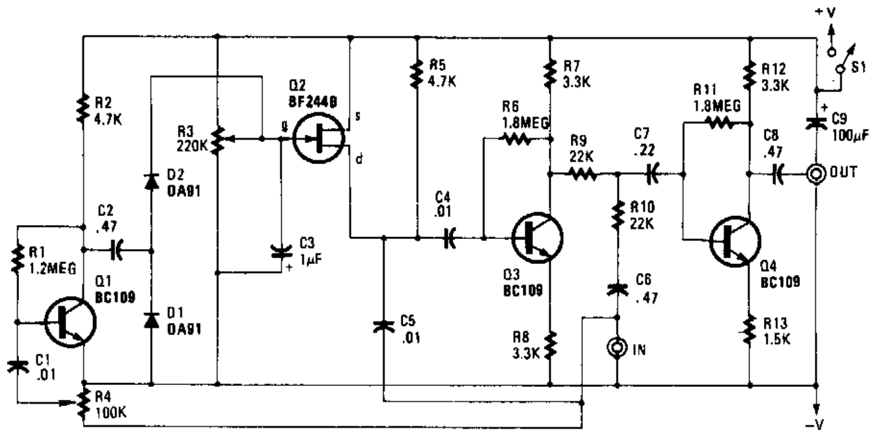
In the state-variable filter circuit, one amplifier performs a summing function and the other two act as integrators. The choice of passive component values is arbitrary, but must be consistent with the amplifier operating range and input signal characteristics. The values shown for C1, C2, R4, R5, and R6 are arbitrary. Preselecting their values will simplify the filter tuning procedures, but other values can be used if necessary.

SCRATCH FILTER



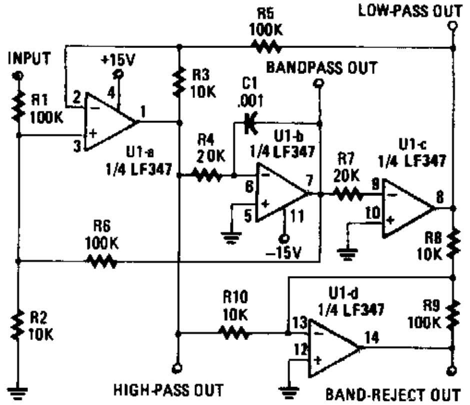
POPULAR ELECTRONICS *Fig. 31-9*

DYNAMIC NOISE FILTER



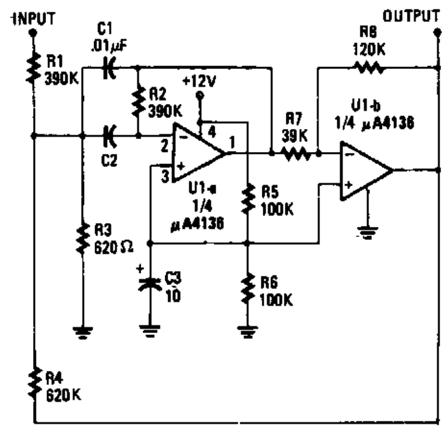
POPULAR ELECTRONICS *Fig. 31-10*

STATE-VARIABLE FILTER WITH MULTIPLE FILTERING OUTPUTS



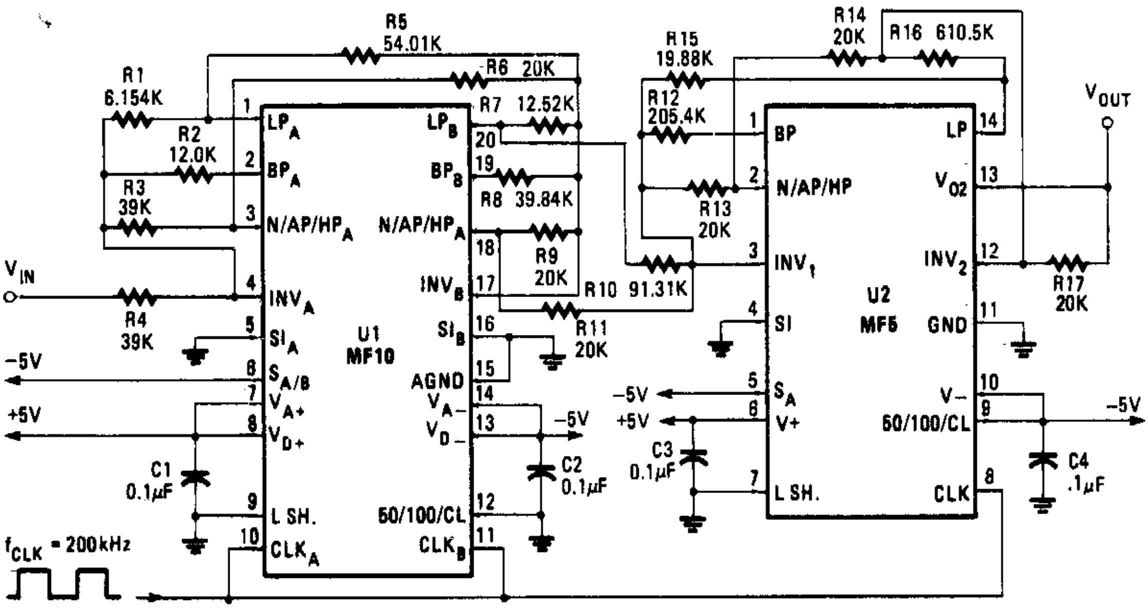
POPULAR ELECTRONICS *Fig. 31-11*

TYPICAL ACTIVE BANDPASS FILTER



POPULAR ELECTRONICS *Fig. 31-12*

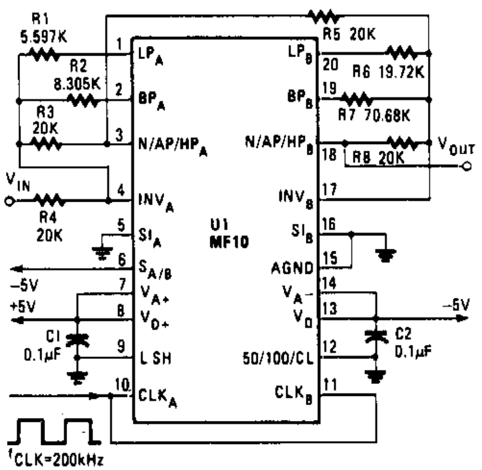
SIXTH-ORDER ELLIPTIC HIGH-PASS FILTER



POPULAR ELECTRONICS

Fig. 31-13

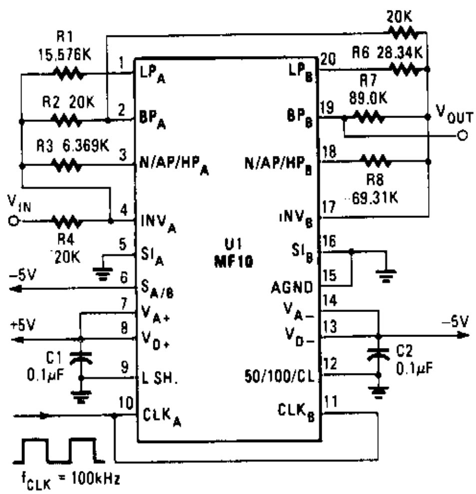
FOURTH-ORDER CHEBYSHEV HIGH-PASS FILTER



POPULAR ELECTRONICS

Fig. 31-14

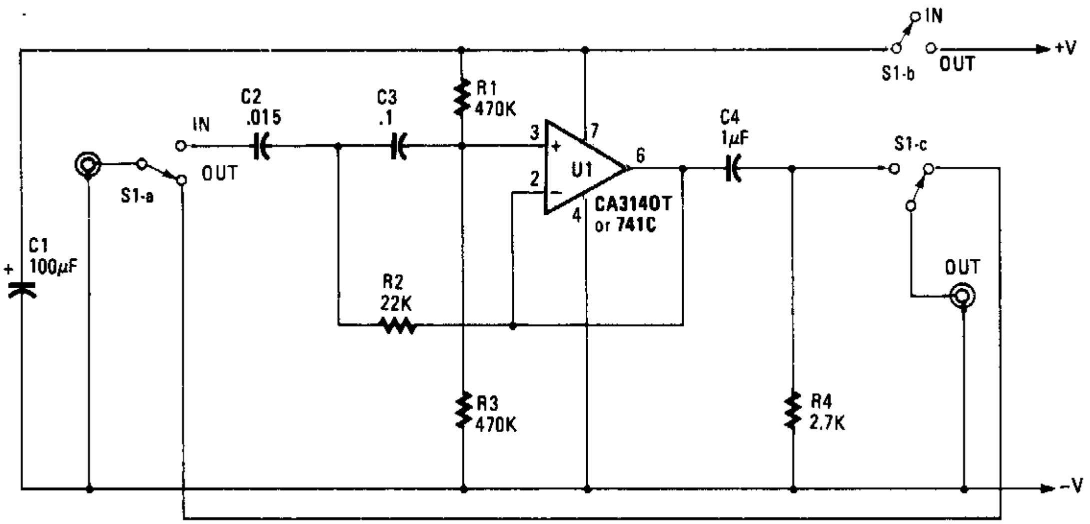
FOURTH-ORDER CHEBYSHEV BANDPASS FILTER



POPULAR ELECTRONICS

Fig. 31-15

RUMBLE FILTER



POPULAR ELECTRONICS

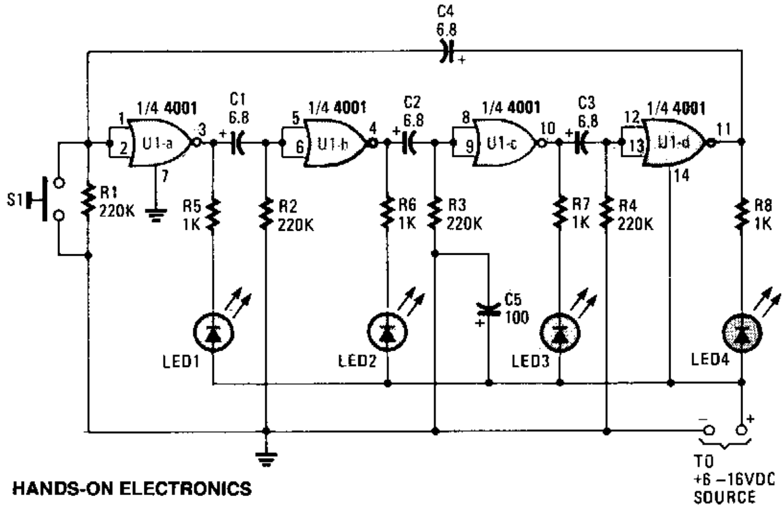
Fig. 31-16

32

Flashers and Blinkers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

RING-AROUND LED FLASHER

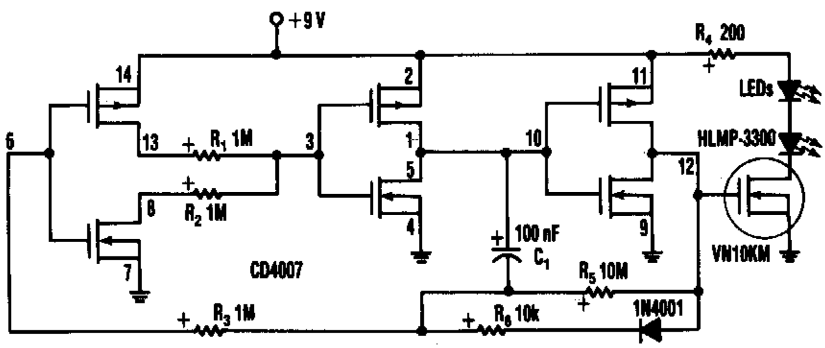


HANDS-ON ELECTRONICS

Fig. 32-1

When power is first turned on, two of the LEDs are on and the other two remain off until the timing cycle reverses. The LEDs flash in pairs, but by pressing and holding S1 closed until only one of the LEDs is on, and then releasing it, the four LEDs can be made to flash in sequential order. The number of LEDs flashing in a sequential ring can be easily increased to eight by adding another 4001 quad NOR gate. Just repeat the circuit and connect the additional circuit in series with the first—input to output—as an extension of the first circuit. When power is connected to the eight-LED flasher circuit, four LEDs will turn on at once and then flash off as the four remaining LEDs come on. As before, just press S1 and hold closed until all but one LED turns off; then the LEDs will begin their sequential march in a circle. You can connect as many circuits in series as you like.

THREE-YEAR LED FLASHER

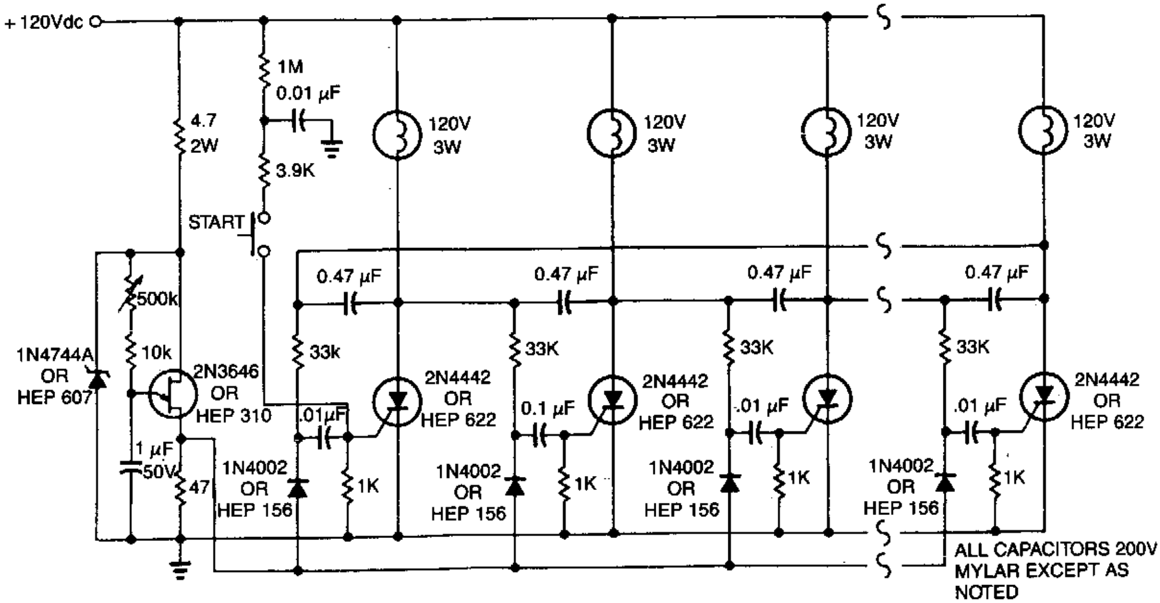


IDEAS FOR DESIGN

Fig. 32-2

Inserting two 1-MΩ resistors, R1 and R2, in the output stage of one of the circuit's inverters limits the current needed by the oscillator to no more than a few μA. This circuit includes a CD4007 package, which has three CMOS inverters. It forms a standard three-inverter oscillator. Resistors R1 and R2, in series with separate drains on inverter pins 8 and 13, limit the oscillator's supply current. Capacitor C1 and resistors R5 set the off time of the oscillator, C1; R6 sets the on time. A VN10KM small-power FET, current-limited by R4, drives two HLMP-3300 LEDs. The LEDs consume about 20 mA for 1 ms. Their average current determines battery life. Since the LEDs in the circuit flash at 1 Hz, the average current drain is about 1/1000 of 20 mA, or 20 μA. A 9-V battery should last about three years at the current drain—essentially the shelf life of an alkaline battery.

SCR RING COUNTER

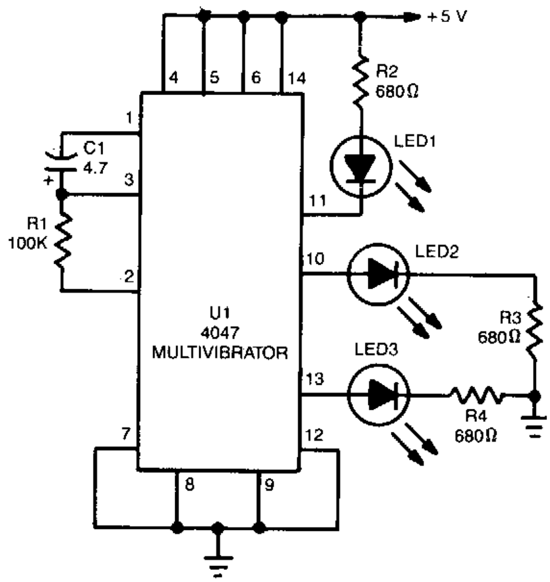


EDN

Fig. 32-3

One lamp at a time is lit in the string to give the appearance of a moving point of light.

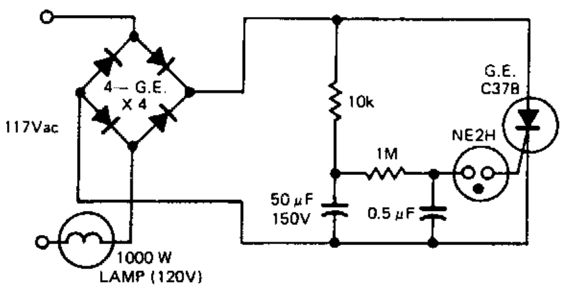
ASTABLE MULTIVIBRATOR



The 4047 is configured as a free-running, astable-multivibrator (oscillator) circuit. That configuration, offers three different outputs. The output pulses at the Q and $\overline{Q}$ output (pins 10 and 11, respectively) are the same as in the previous two circuits. The third output at pin 13 pulses twice as often as the outputs at 10 and 11. So, the circuit can be used to simultaneously provide both positive- and negative-trigger signals since the Q and $\overline{Q}$ output are never in the same state, and a clock frequency. Thus, the 4047 can replace both a simple oscillator (the 555, for instance) and a flip-flop in some applications.

POPULAR ELECTRONICS *Fig. 32-4*

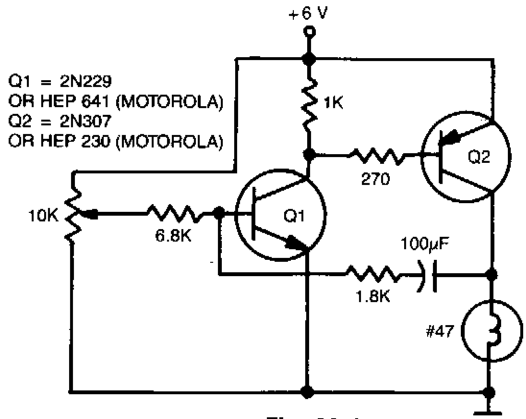
AC FLASHER



EDN *Fig. 32-5*

This ac line-operated flasher uses an SCR and is capable of flashing a large lamp. Flashing rate is determined by the 10-KΩ resistor and the 50-mF capacitor. Increasing or decreasing the value of the capacitor has a corresponding effect on the flash rate.

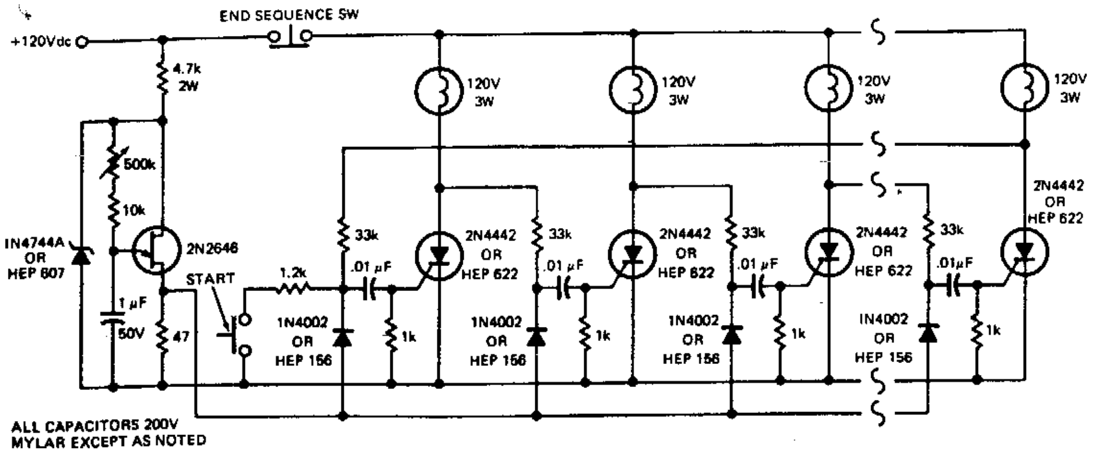
SINGLE-LAMP FLASHER



EDN *Fig. 32-6*

The flash rate is controlled by a complementary multivibrator consisting of an npn and a pnp transistor.

SCR CHASER

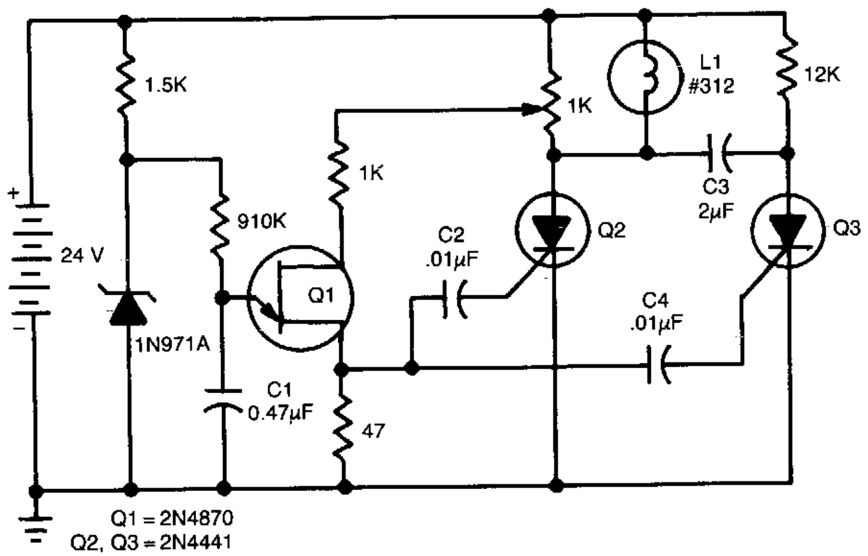


EDN

Fig. 32-7

Each lamp lights in succession to give the appearance of a growing column.

SCR FLASHER

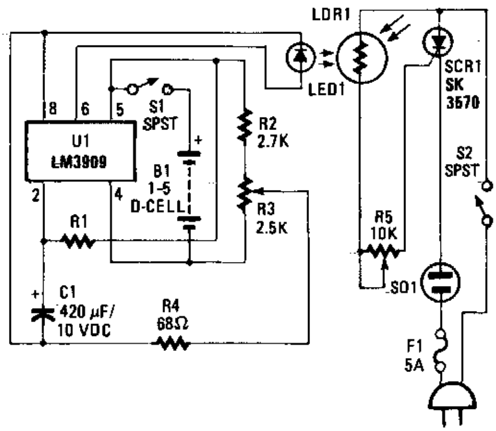


EDN

Fig. 32-8

This dc flasher uses two SCRs and a unijunction oscillator clock to set the flash rate, which can be varied by changing the value of C1.

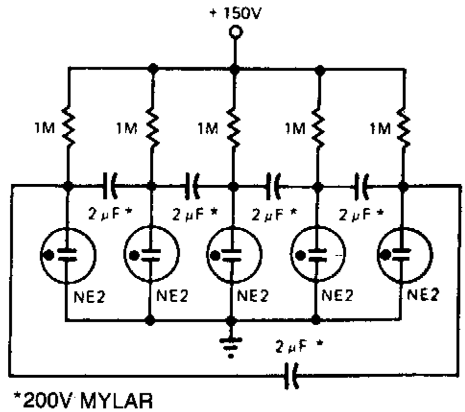
INCANDESCENT LIGHT FLASHER



POPULAR ELECTRONICS

Fig. 32-9

FIVE-LAMP NEON FLASHER



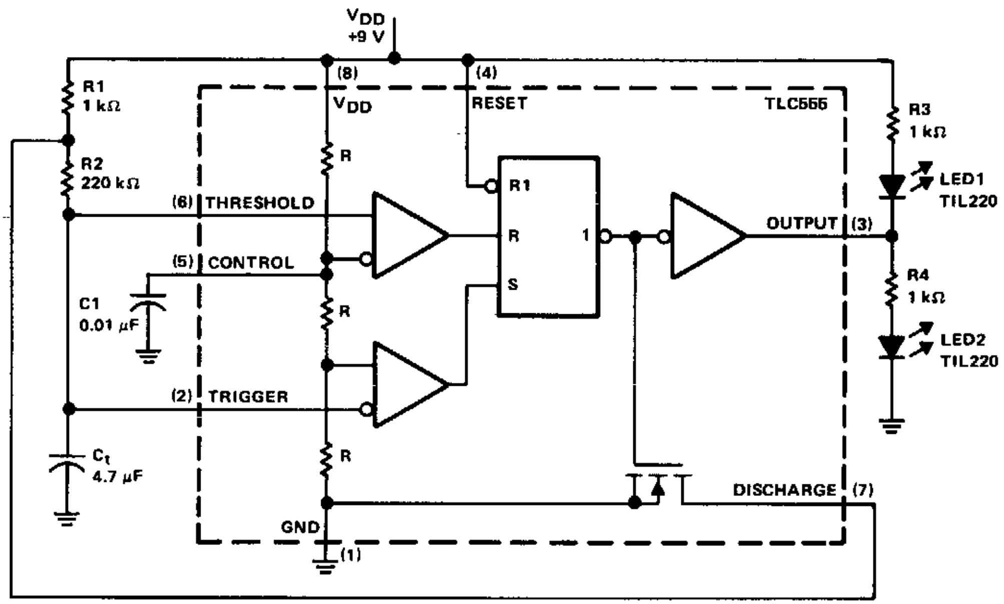
*200V MYLAR

EDN

Fig. 32-10

In this circuit, the number of lamps can be increased almost without limit.

ALTERNATING LED FLASHER



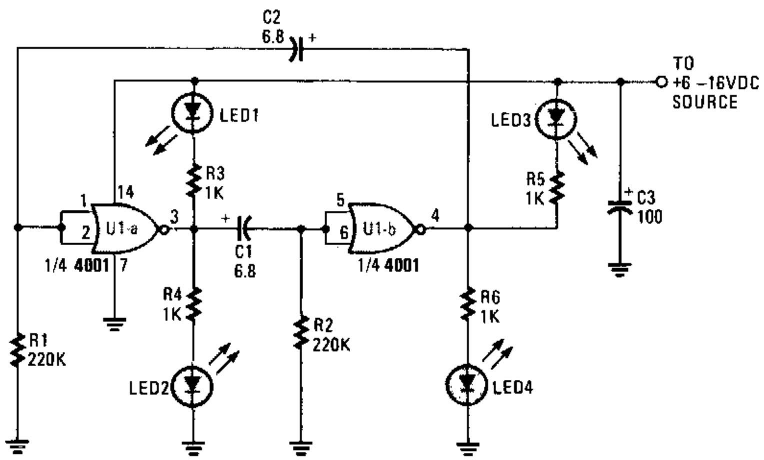
Reprinted by permission of Texas Instruments.

Fig. 32-11

The timing components are R1, R2, and C_t. C1 is a bypass capacitor used to reduce the effects of noise. At start-up, the voltage across C_t is less than the trigger level voltage ($1/3\ V_{DD}$), causing the timer to be triggered via pin 2. The output of the timer at pin 3 increases, turning LED1 off, LED2 on, the discharge transistor at pin 7 off, and allowing C_t to charge through resistors R1 and R2. When capacitor C_t charges to the upper threshold voltage ($2/3\ V_{DD}$), the flip-flop is reset and the output at pin 3 decreases. LED1 is turned on, LED2 is turned off, and capacitor C_t discharges through resistor R2 and the discharge transistor. When the voltage at pin 2 reaches $1/3\ V_{DD}$, the lower threshold or trigger level, the timer triggers again and the cycle is repeated.

The totem-pole output at pin 3 is a square wave with a duty cycle of about 50%. The output alternately turns on each LED at slightly less than one blink per second. If the unit is battery operated, the TLC555 uses minimum current to produce this function. With a 9-V battery, the circuit draws 5 mA (no load) and 15 mA when turning on an LED. Most of the on current is for the LED.

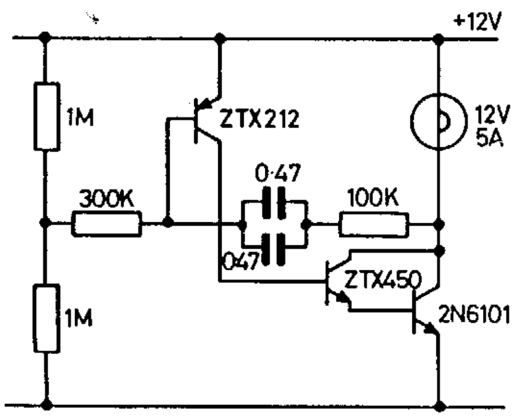
CMOS FLASHER



HANDS-ON ELECTRONICS *Fig. 32-12*

Uses a low cost CMOS IC to turn four LEDs on and off at a rate that is set by the values of R1, R2, C1, and C2. The pulse rate for the component values given for R1 and R2 is about one cycle every four seconds. By lowering the values of R1 and R2 to 220 KΩ, the pulse rate increases to 1 Hz. The LEDs flash in pairs, with LED1 and LED4 turning together for one half of the time period, while LED2 and LED3 are on for the other half. The on/off duration of each pair of LEDs can be increased or decreased by changing the value of one of coupling capacitors C1 or C2. Increasing either capacitor's value by a factor of 10 will also increase the ON time of a pair of the LEDs for about the same factor.

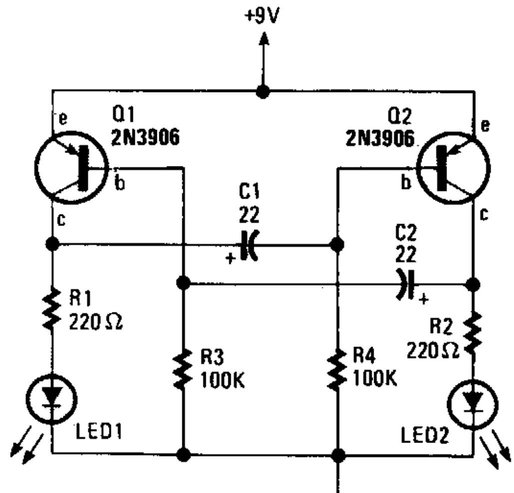
60-W FLASHING LIGHT



ZETEX, formerly FERRANTI *Fig. 32-13*

The 2N6101 transistor should be mounted on a small heatsink. The 300-KΩ resistor controls the off period and might need to be adjusted if transistor gains are high. The 100-KΩ resistor controls the on period.

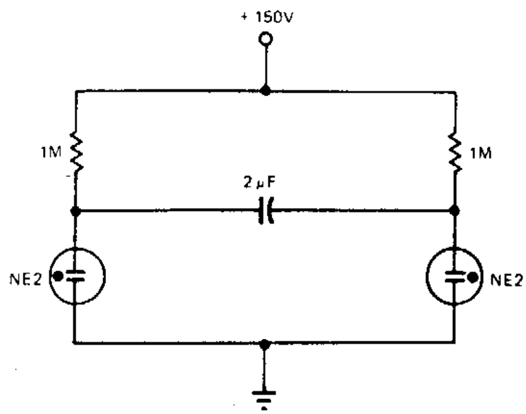
ALTERNATING LED FLASHER



POPULAR ELECTRONICS *Fig. 32-15*

The alternating LED flasher is simply a two-transistor oscillator with LEDs connected to the collector of each transistor, so that they light in time with the circuit's oscillations.

TWO-STATE NEON OSCILLATOR

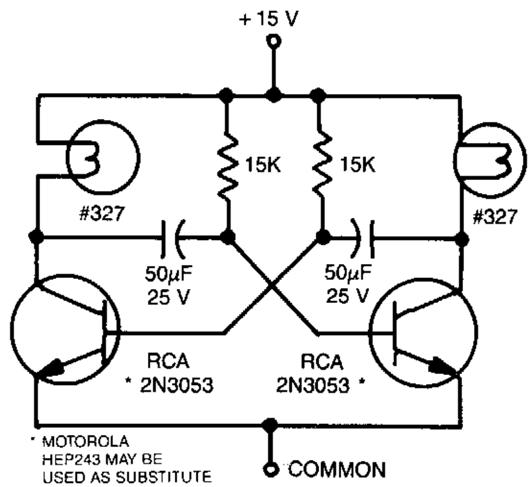


* 200V MYLAR

EDN *Fig. 32-14*

The number of lamps is easily increased in this oscillator.

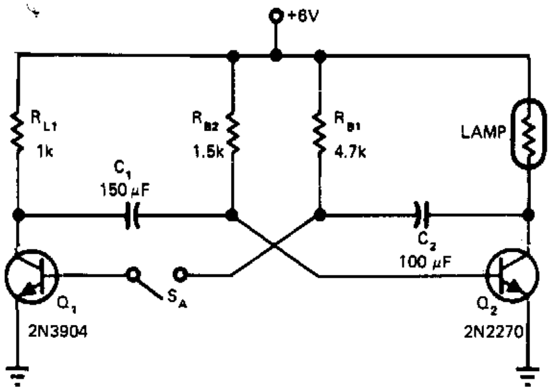
TRANSISTOR FLASHER



EDN *Fig. 32-16*

This astable multivibrator uses incandescent lamps in place of collector load resistors. The lamps flash on and off alternately.

MINIMUM COMPONENT FLASHER



ELECTRONIC DESIGN

Fig. 32-17

Opening S_A, changes the indicator lamp from flashing to steady-lit condition. The 6-V incandescent lamp on the collector of Q2 requires about 0.3 A. A 1-KΩ load resistor limits Q1's collector current to about 6 mA. The circuit is, therefore, asymmetrical with respect to the on currents of the transistors, allowing use of a much smaller transistor for Q1 than for Q2.

LAMP FLASHER

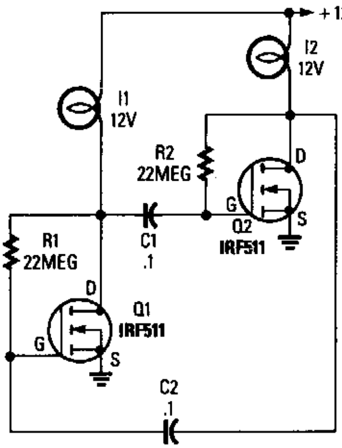


Fig. 32-18

POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

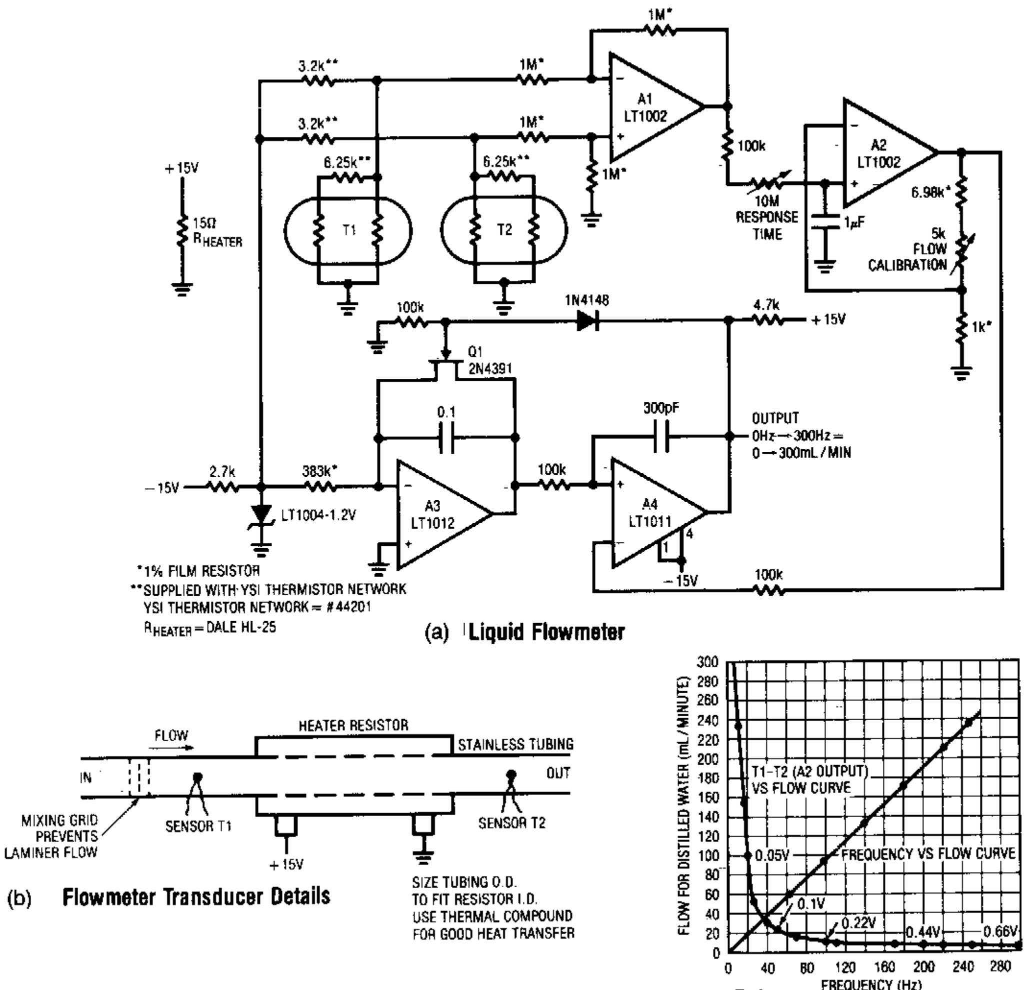
The circuit is built around two power FETs, which are configured as a simple astable multivibrator to alternately switch the two lamps on and off. The rc values given sets the flash rate to about 1/3 Hz. By varying either the resistor or capacitor values, almost any flash rate can be obtained. Increase either C1 and C2, or R1 and R2, and the flash rate slows. Decrease them and the rate increases.

33

Flow Detector

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

LOW FLOW-RATE THERMAL FLOWMETER



(a) Liquid Flowmeter

(b) Flowmeter Transducer Details

(c) Flowmeter Response Data

Fig. 33-1

LINEAR TECHNOLOGY CORP.

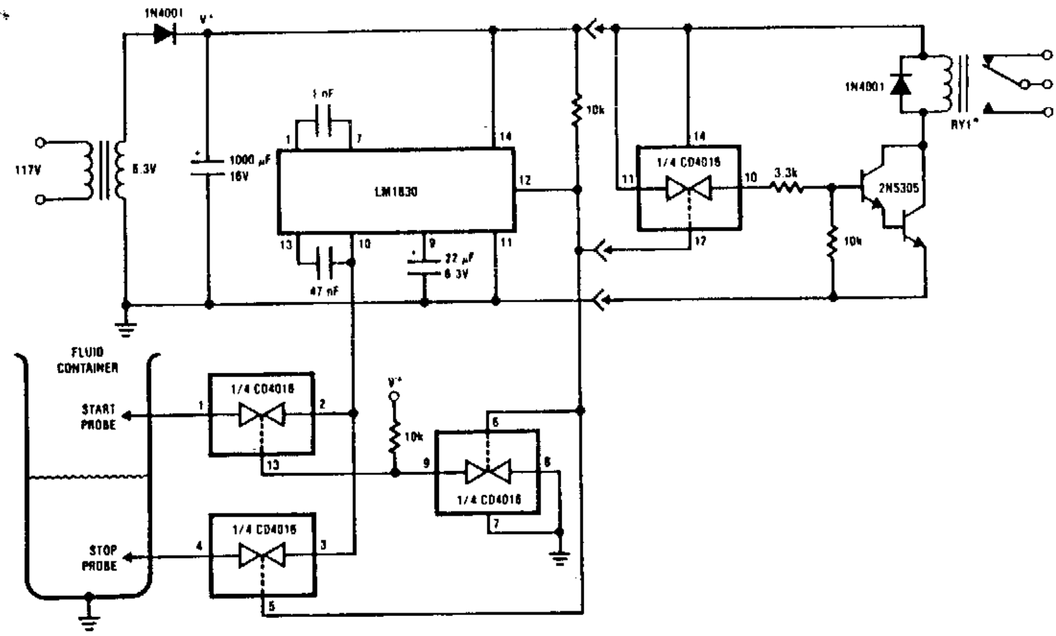
This design measures the differential temperature between two sensors. Sensor T1, located before the heater resistor, assumes the fluid's temperature before it is heated by the resistor. Sensor T2 picks up the temperature rise induced into the fluid by the resistor's heating. The sensor's difference signal appears at Al's output. A2 amplifies this difference with a time constant set by the 10 MΩ adjustment. Fig. 33-1c shows A2's output versus flow rate. The function has an inverse relationship. A3 and A4 linearize this relationship, while simultaneously providing a frequency output. A3 functions as an integrator that is biased from the LT1004 and the 338-KΩ input resistor. Its output is compared to A2's output at A4. Large inputs from A2 force the integrator to run for a long time before A4 can increase, turning on Q1 and resetting A3. For small inputs from A2, A3 does not have to integrate long before resetting action occurs. Thus, the configuration oscillates at a frequency which is inversely proportional to A2's output voltage. Since this voltage is inversely related to flow rate, the oscillation frequency linearly corresponds to flow rate.

34

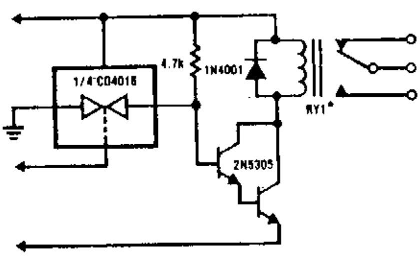
Fluid and Moisture Detectors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

FLUID-LEVEL CONTROL



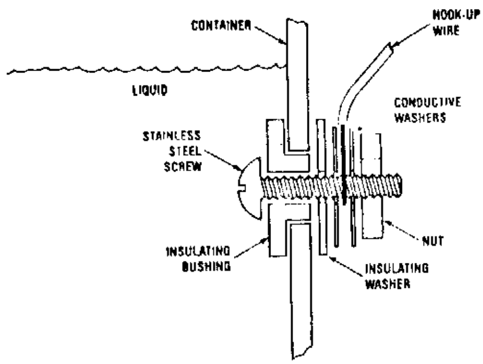
"Emptying" Processes are Controlled with this Circuit



*RY1 = Magnecraft Part # W388CQX-5

FIGURE 1(b). Filling Processes are Implemented with this Output Circuit and Relabeled Probes

NATIONAL SEMICONDUCTOR CORP.

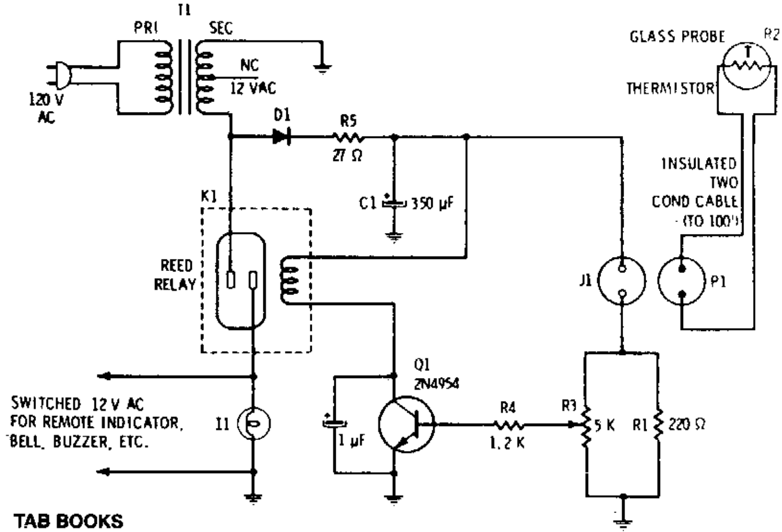


A sealing compound applied externally protects hook-up wire and prevents leaks.

Typical Probe Installation ***Fig. 34-1***

This circuit is designed to detect the presence or absence of aqueous fluids. An ac signal generated on-chip is passed through two probes within the fluid. A detector determines the presence of the fluid by using the probes in a voltage divider circuit and measuring the signal level across the probes. An ac signal is used to prevent plating or dissolving of the probes as occurs when a dc signal is used. A pin is available for connecting an external resistance in cases where the fluid impedance is not compatible with the internal 13-KΩ divider resistance.

FLOOD ALARM OR TEMPERATURE MONITOR

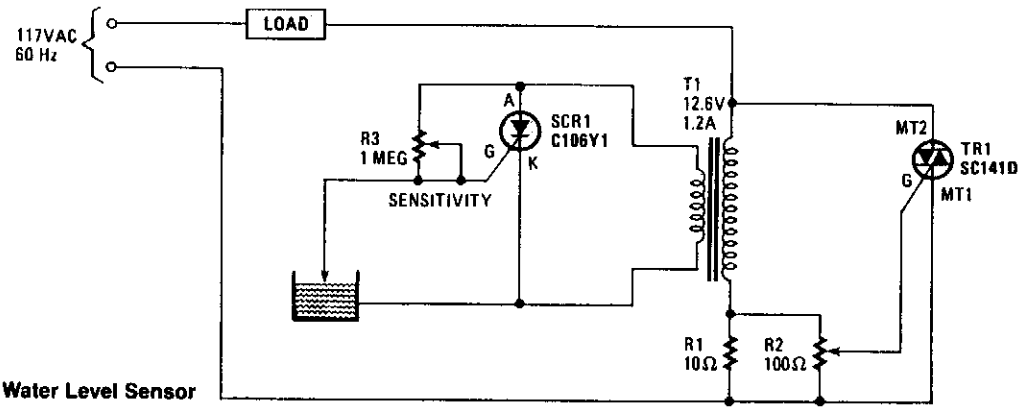


TAB BOOKS

Fig. 34-2

Filtered 15 Vdc is applied to a series circuit consisting of thermistor R2 and parallel combination of resistors R1 and R3. Transistor Q1 acts as a switch whose state is determined by the setting of potentiometer R3, which is first set so just enough current flows into the base to switch on when the thermistor is in contact with air. When the resistance of the thermistor decreases, the voltage at the base of Q1 rises. When the base current reaches the preset level, the transistor conducts and passes current through the reed relay coil, closing the reed relay contacts. Current at the base of transistor Q1 is determined by the environment into which the termistor is inserted.

WATER-LEVEL SENSOR AND CONTROL



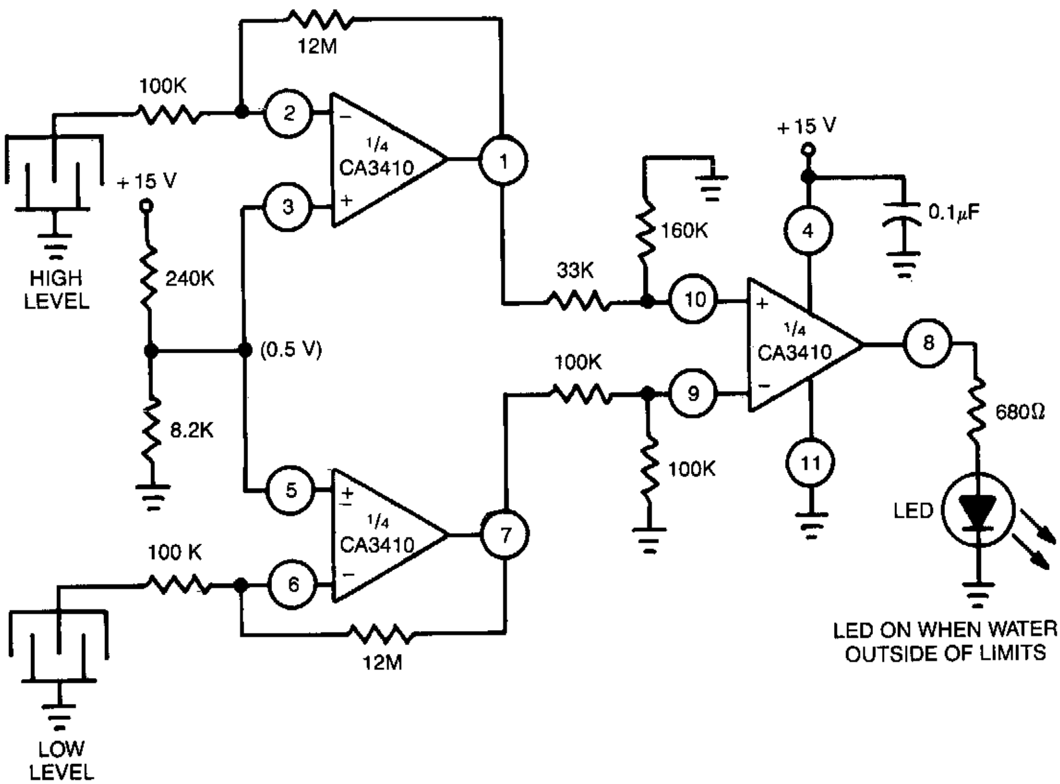
Water Level Sensor

HANDS-ON ELECTRONICS

Fig. 34-3

When the water level is low, the probe is out of the water and SCR1 is triggered on. It conducts and imposes a heavy load on transformer T1's secondary winding. That load is reflected back into the primary, gating triac TR1 on, which energizes the load. If the load is an electric valve in the water-supply line, it will open and remain open until the water rises and touches the probe; this shorts SCR1's gate and cathode, thereby turning off the SCR1, which effectively open-circuits the secondary. That open-circuit condition, when reflected back to the primary winding, removes the triac's trigger signal, thereby turning the water off.

DUAL LIQUID-LEVEL DETECTOR



GE/RCA

Fig. 34-4

Uses CA3410 quad BiMOS op amp to sense small currents. Because the op amp's input current is low, a current of only 1 μA passing through the sensor will change the converter's output by as much as 10 to 12 V.

SOIL-MOISTURE METER

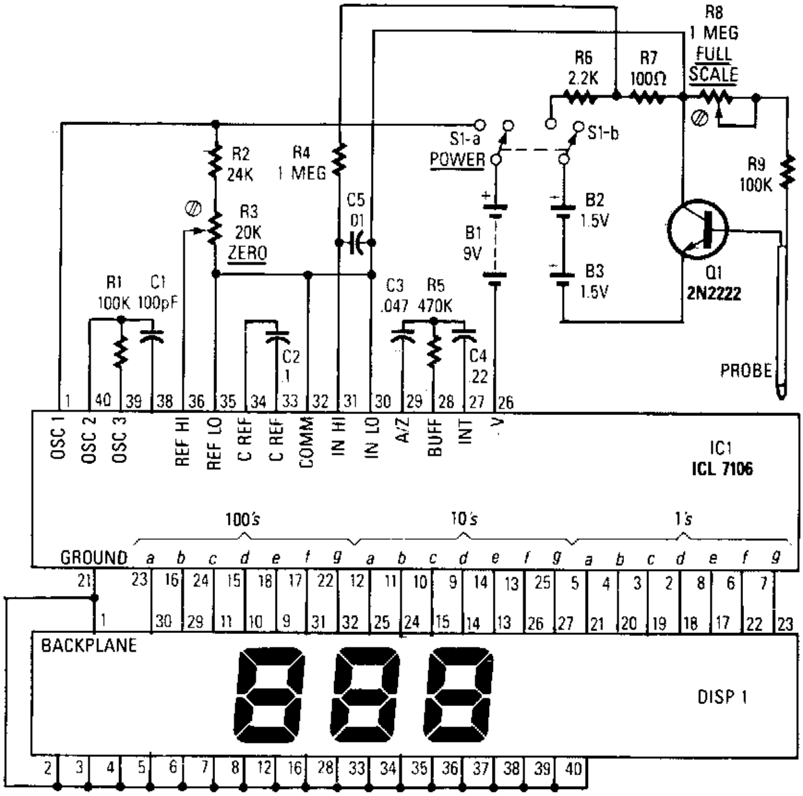


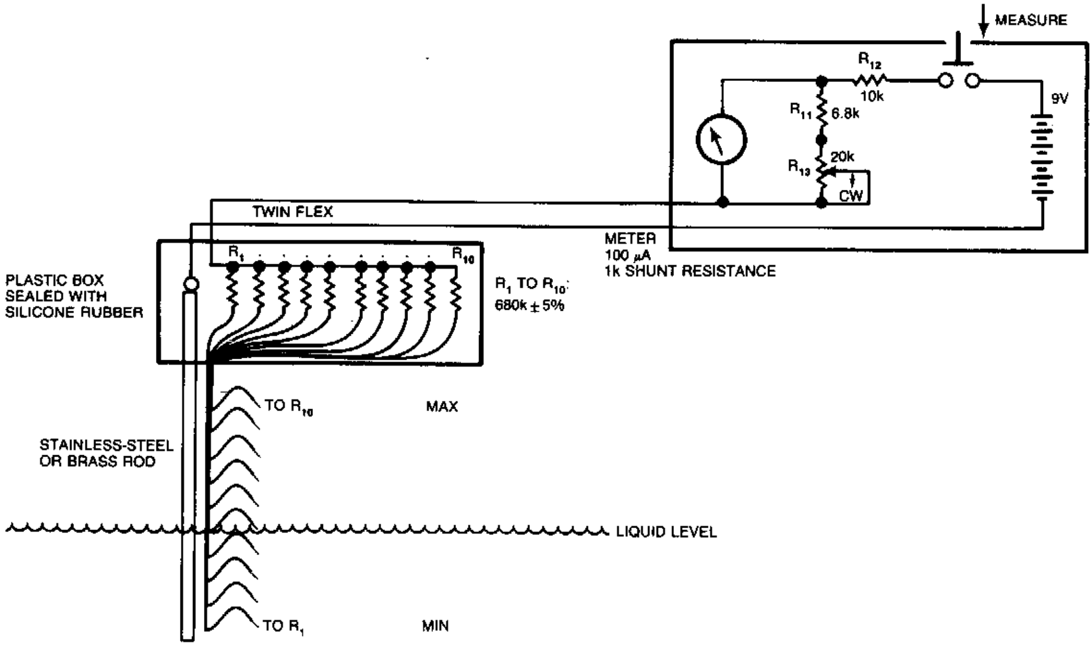
Fig. 34-5

IC1, an Intersil ICL7106, contains an a/d converter, a $3^1/_2$-digit LCD driver, a clock, a voltage reference, seven segment decoders, and display drivers. A similar part, the ICL7107, can be used to drive seven segment LEDs. The probe body is a five-inch length of light-weight aluminum tubing. The leads from the circuit are connected to the body and tip of the probe. The sensor functions as a variable resistor that varies Q1's base current, hence its collector current. The varying collector current produces a varying voltage across 100 Ω resistor R7, and that voltage is what IC1 converts for display.

The LCD consumes about 25 μA, and IC1 consumes under 2 mA, so the circuit will run for a long time when it is powered by a standard 9-V battery. Current drain of the two 1.5-V AA cells is also very low: under 300 μA.

To calibrate, rotate R3 to the center of its range. Then place the end of the probe into a glass of water and adjust R8 for a reading of 100. When you remove the probe from the water, the LCD should indicate 000. You might have to adjust R3 slightly for the display to indicate 000. If so, readjust R8 with the probe immersed. Check for a reading of 000 again with probe out of water.

LIQUID-LEVEL CHECKER

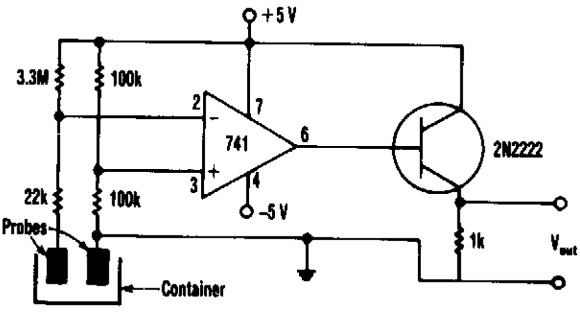


EDN

Fig. 34-6

Although many circuits use the varying-capacitance method for checking liquid levels, this simple resistive circuit is much easier to construct. Even a tank of a liquid, such as water, has sufficient conductive salts in solution for this method to work. The probe uses a metal rod that supports 10 insulated wires, which have stripped ends pointing down. As the level of liquid rises, resistors R1 through R10 are successively brought into circuit, each drawing an extra 10 μA through the meter. Shunt resistors R11 and R13 calibrate the meter for a full-scale reading when the tank is full. Resistor R12 limits the current through the meter. If tank isn't rectangular—ie, if the volume of the liquid it contains isn't directly proportional to the liquid's depth—space the resistors accordingly or use a nonlinear progression of resistor values and retain constant resistor spacing.

LIQUID-LEVEL MONITOR



ELECTRONIC DESIGN *Fig. 34-7*

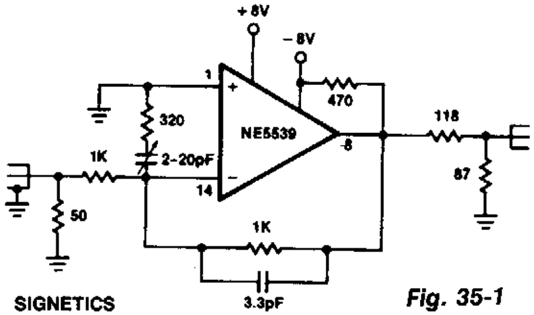
This monitor uses a common 741 amp configured as a comparator and a low cost nontransistor as an output driver. With no liquid detected, a voltage of about 2.92 V is present in the op amp's inverting input at pin 2. The 100-KΩ resistors establish a reference voltage of +2.5 V at the noninverting input at pin 3 of the op amp. Under those conditions, the op amp's output is −3.56 V, which keeps the 2N2222 transistor turned off and the voltage across its 1-KΩ output load resistor at 0 V. When liquid reaches the probes, the 3.3-MΩ and 22-KΩ resistor circuit conductively connects to ground. When enough current, about 1.4 μA, flows through the liquid, the small 30 mV drop developed across the 22-KΩ resistor drives the op amp to deliver an output voltage of about 4.42 V. This voltage then drives a 2N2222 transistor into saturation, which generates a voltage drop of about 3.86 V across its 1-KΩ output load resistor.

35

Followers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

HIGH-FREQUENCY INVERTING FOLLOWER



SIGNETICS

Fig. 35-1

HIGH-FREQUENCY NONINVERTING FOLLOWER

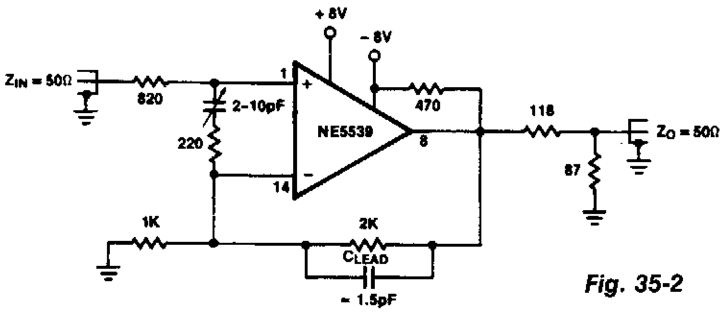
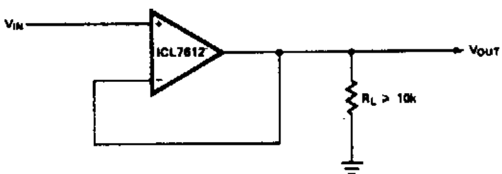


Fig. 35-2

SIGNETICS

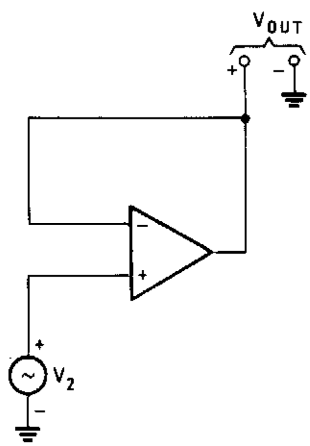
SIMPLE FOLLOWER



INTERSIL

Fig. 35-3

VOLTAGE FOLLOWER



HANDS-ON ELECTRONICS

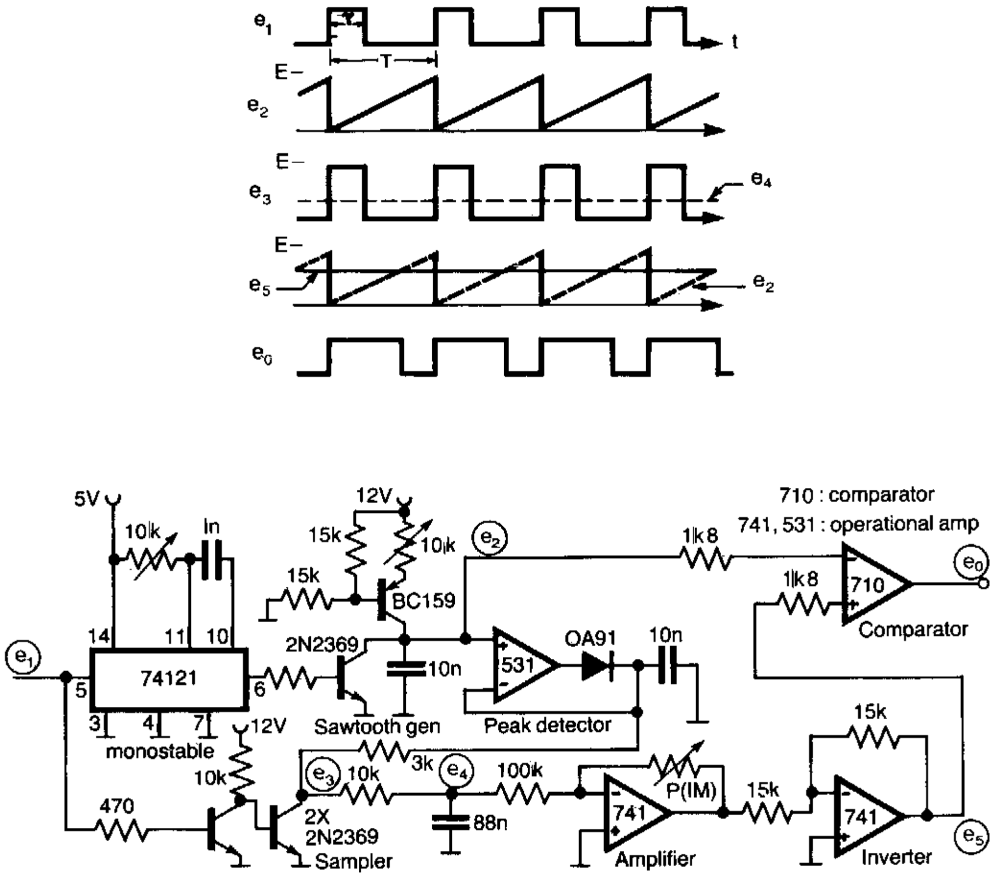
Fig. 35-4

36

Frequency Multipliers and Dividers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PULSE-WIDTH MULTIPLIER

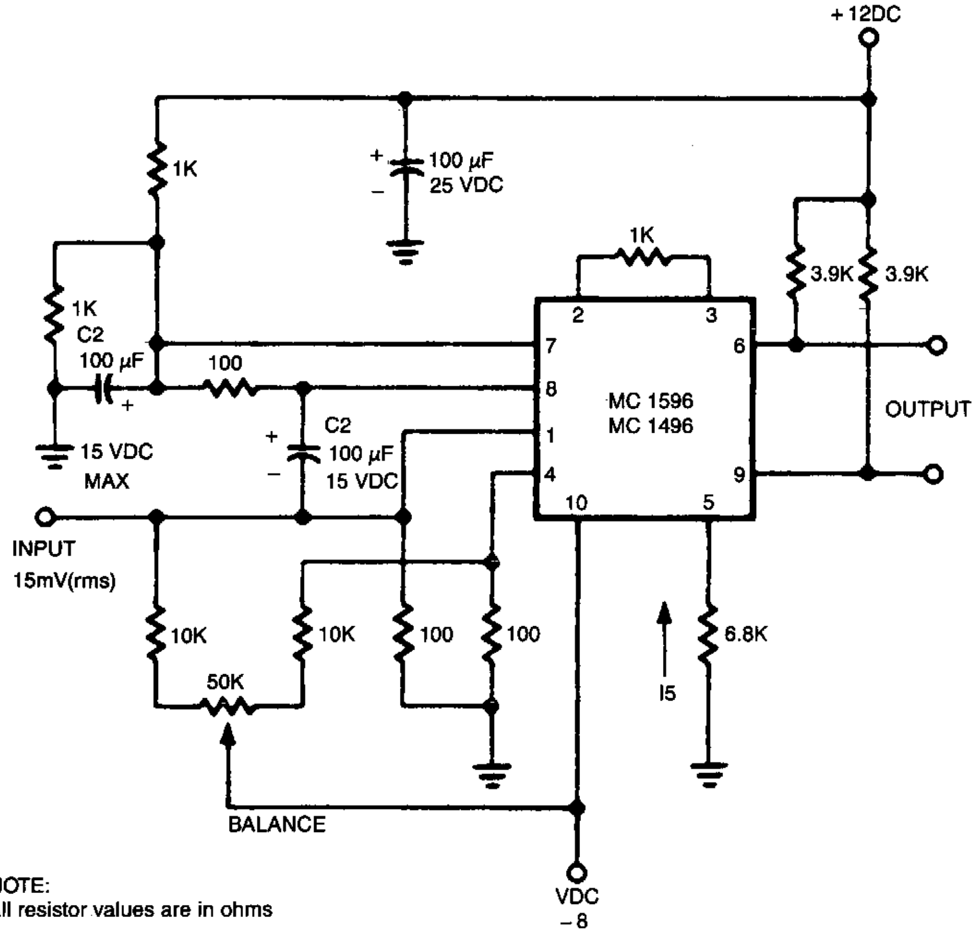


ELECTRONIC ENGINEERING

Fig. 36-1

This circuit for multiplying the width of incoming pulses by a factor greater or less than unity is simple to build and has the feature that the multiplying factor can be selected by adjusting one potentiometer only. The multiplying factor is determined by setting the potentiometer in the feedback of the 741 amplifier. The input pulses e_1, width T and repetition period T is used to trigger a sawtooth generator at its rising edges to produce the waveform e_2 having a peak value of E volts. This peak value is then sampled by the input pulses to generate the pulse train e_3 having an average value of e_4 ($=E_T/T$) which is proportional to T and independent on T. The dc voltage e_4 is amplified by a factor k and compared with sawtooth waveform e_2 giving output pulses of duration k_T. The circuit is capable of operating over the frequency range 10 kHz – 100 kHz. Note that k should be chosen less than T/T to ensure accurate pulse-width multiplication.

FREQUENCY DOUBLER



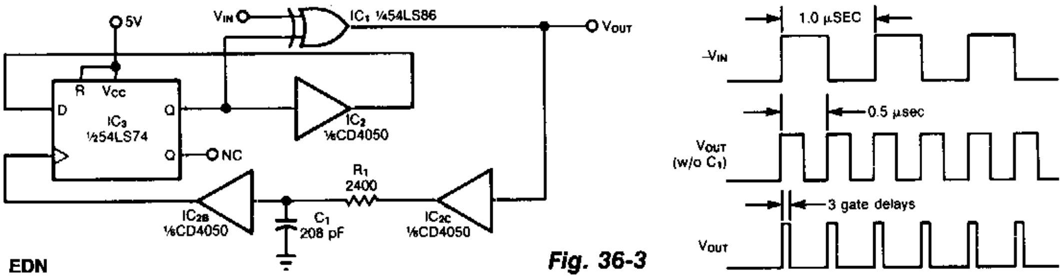
NOTE:
All resistor values are in ohms

SIGNETICS

Fig. 36-2

The output contains the sum component, which is twice the frequency of the input, since both input signals are the same frequency.

DIGITAL FREQUENCY DOUBLER



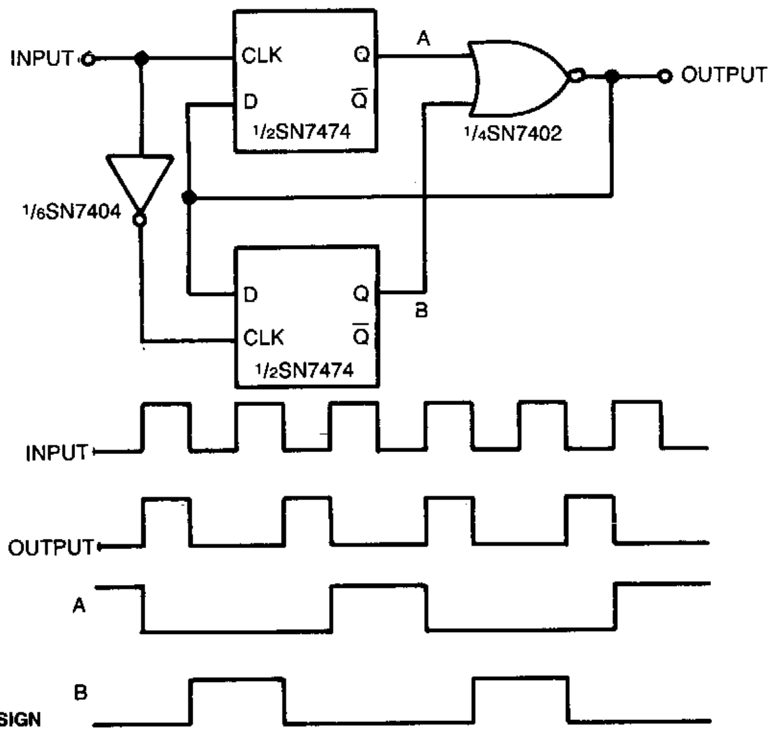
EDN

Fig. 36-3

The circuit doubles the frequency of a digital signal by operating on both signal edges. Each transition causes exclusive-OR gate IC1 to produce a pulse, which clocks flip-flop IC3 after propagating through buffers IC2C and IC2B. If you remove capacitor C1, the circuit produces narrow output pulses. By including C1, you can obtain a desired duty cycle for a given input frequency f_{IN}. The C1 value for an approximate 50% duty cycle is:

$$C1 = \frac{1}{2R_1 f_{IN}}$$

DIVIDE-BY-1 1/2 CIRCUIT

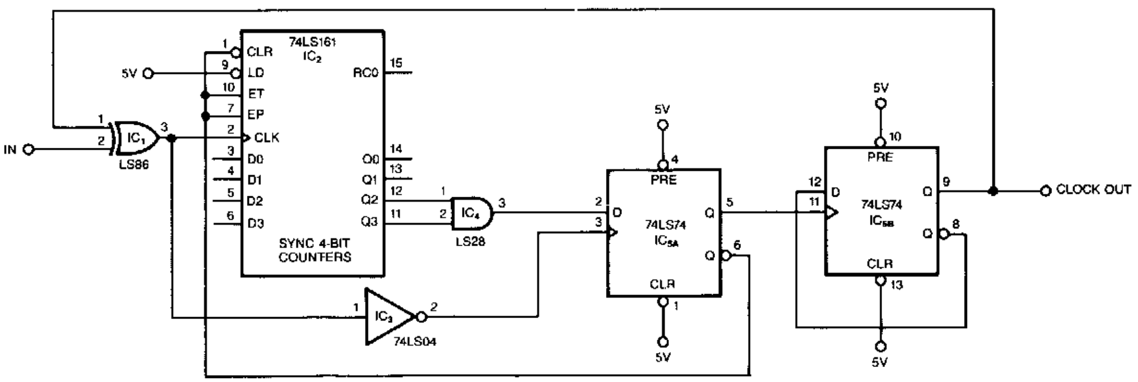


ELECTRONIC DESIGN

Fig. 36-4

An input signal drives both SN7474 D-type flip-flops, which are positive edge-triggered devices. A low-to-high input signal transition triggers the A flip-flop, while a high-to-low input signal transition triggers the B flip-flop via the SN7404 inverter. Either flip-flop in the high state will cause the output to decrease via the SN7402 NOR gate. This in turn disables the opposite flip-flop from going to the high state. The flip-flop in the high state remains there for one clock period, then it is clocked low. With both flip-flops low, the output increases, enabling the opposite flip-flop to be clocked high one-half clock cycle later. This alternate enabling and disabling action of the flip-flops results in a divide-by-1$^1/_2$ function. That is, three clock pulses in, produce two evenly spaced clock pulses out. The circuit has no lock-up states and no inherent glitches. Replacing the NOR gate with an SN7400 NAND gate inverts the A, B, and output signals. By adding simple binary or BCD counters, counting chains, such as divide-by-3, -6, -12, -24, -15, -30, etc., can be generated using the divide-by-1$^1/_2$ circuit as a basis.

ODD-NUMBER COUNTER DIVIDER

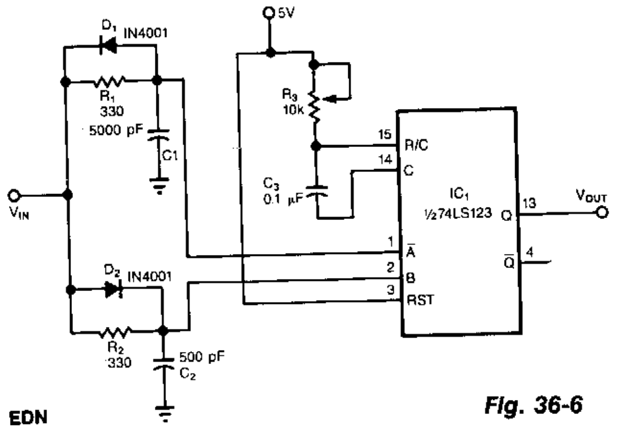


EDN

Fig. 36-5

This circuit, shown symmetrically, divides an input by virtually any odd number. The circuit counts $n + {}^1/_2$ clocks twice to achieve the desired divisor. By selecting the proper n, which is the decoded output of the LS161 counter, you can obtain divisors from 3 to 31. The circuit, as shown, divides by 25; you can obtain higher divisors by cascading additional LS161 counters. The counter and IC5A form the $n + {}^1/_2$ counter. Once the counter reaches the decoded count, n, IC5A ticks off an additional $^1/_2$ clock, which clears the counter and puts it in hold. Additionally, IC5A clocks IC5B, which changes the clock phasing through the XOR gate, IC1. The next edge of the input clocks IC5A, which reenables the counter to start counting for an additional $n + {}^1/_2$ cycles. Although the circuit has been tested at 16 MHz, a worst-case timing analysis reveals that the maximum input frequency is between 7 and 8 MHz.

SINGLE-CHIP FREQUENCY DOUBLER



EDN

Fig. 36-6

The frequency doubler uses only one IC. Like other doublers, this circuit uses both the rising and falling edges of the input signals to produce digital pulses, thus effectively doubling the input's frequency.

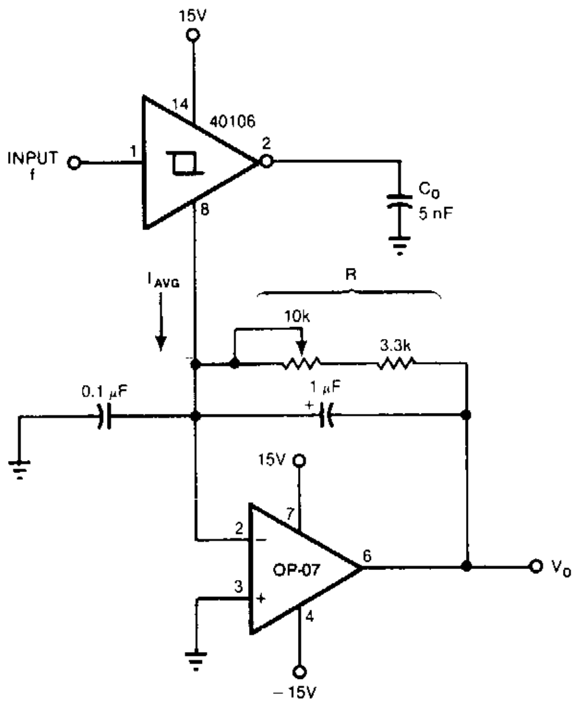
Without the rc networks at IC1 inputs, IC1 would not produce any output pulses. However, the rc networks delay one edge with respect to the other. The A input lags the B input for positive-going edges, and the B input lags the A input for negative-going ones. You can vary the output duty cycle from 0 to 100% by varying R3. IC1's minimum output pulse width defines the maximum frequency of this circuit.

37

Frequency-to-Voltage Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

FREQUENCY-TO-VOLTAGE CONVERTER



EDN

Fig. 37-1

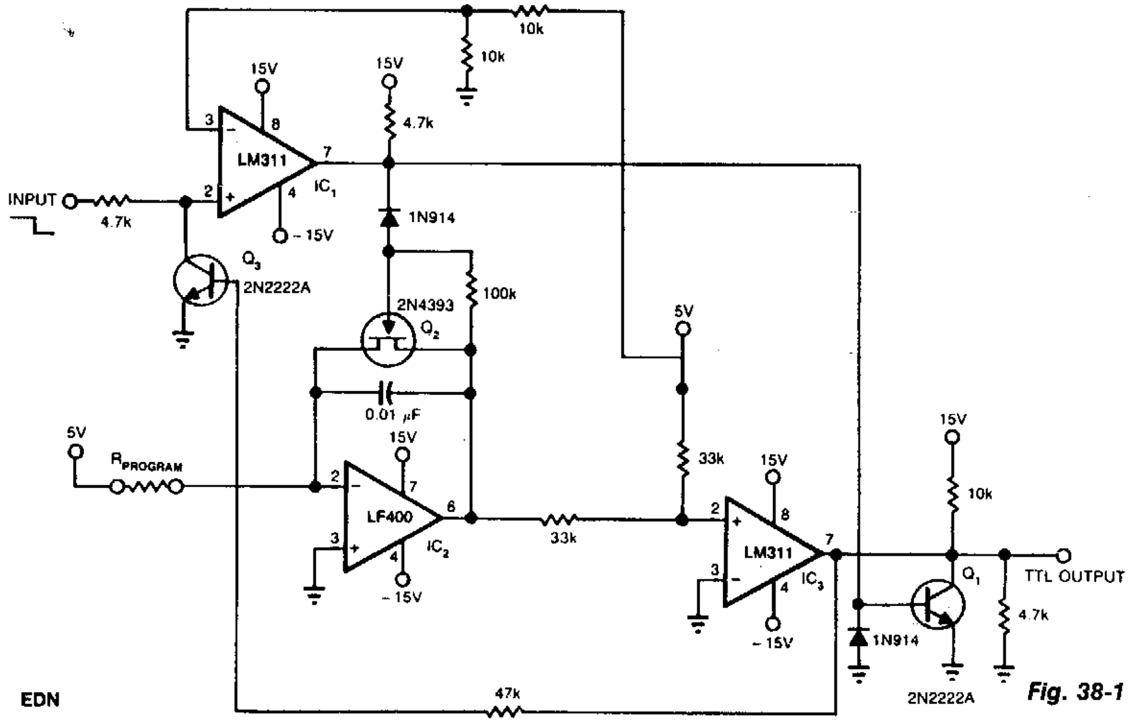
Six components can configure a circuit whose output voltage is proportional to its input frequency. The average current (I_{AVG} from the 40106 Schmitt trigger inverter's ground pin 8 is linearly dependent on the frequency at which $C0$ is discharged into the op amp's summing junction. The op amp forces this current to flow through the 13.33-KΩ feedback resistor, producing a corresponding voltage drop. This frequency-to-voltage converter yields 0 to −10 V output with 0 to 10 kHz input frequencies.

38

Function Generators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PRECISION ONE SHOT

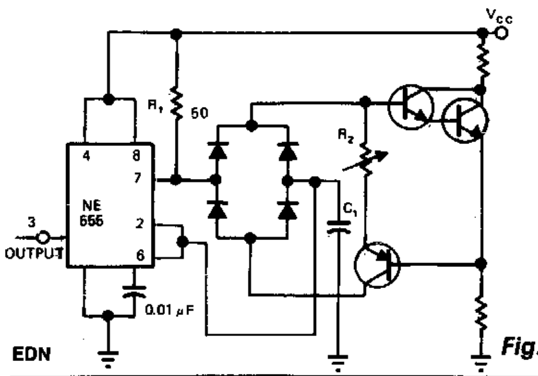


EDN

Fig. 38-1

If you need a wide-range, resistor-programmable monostable multivibrator, you can program the circuit for pulse widths from 1 μs to 10s – 10^7:1 range. A high-to-low transition at the input causes IC1's output to switch low, thereby turning off Q1 and Q2. With the latter transistor turned off, IC3's output increases and the output of IC2 begins to ramp toward the negative supply level at a rate determined by the 0.01-μF capacitor and the programming resistor. When IC2's output voltage reaches – 5 V, IC3's output switches low. If you anticipate input pulses shorter than the desired output pulses, Q3 is necessary. This transistor keeps IC1s input low while an output pulse is present, thereby preventing inadvertent resetting of the one shot.

LINEAR TRIANGLE-WAVE TIMER

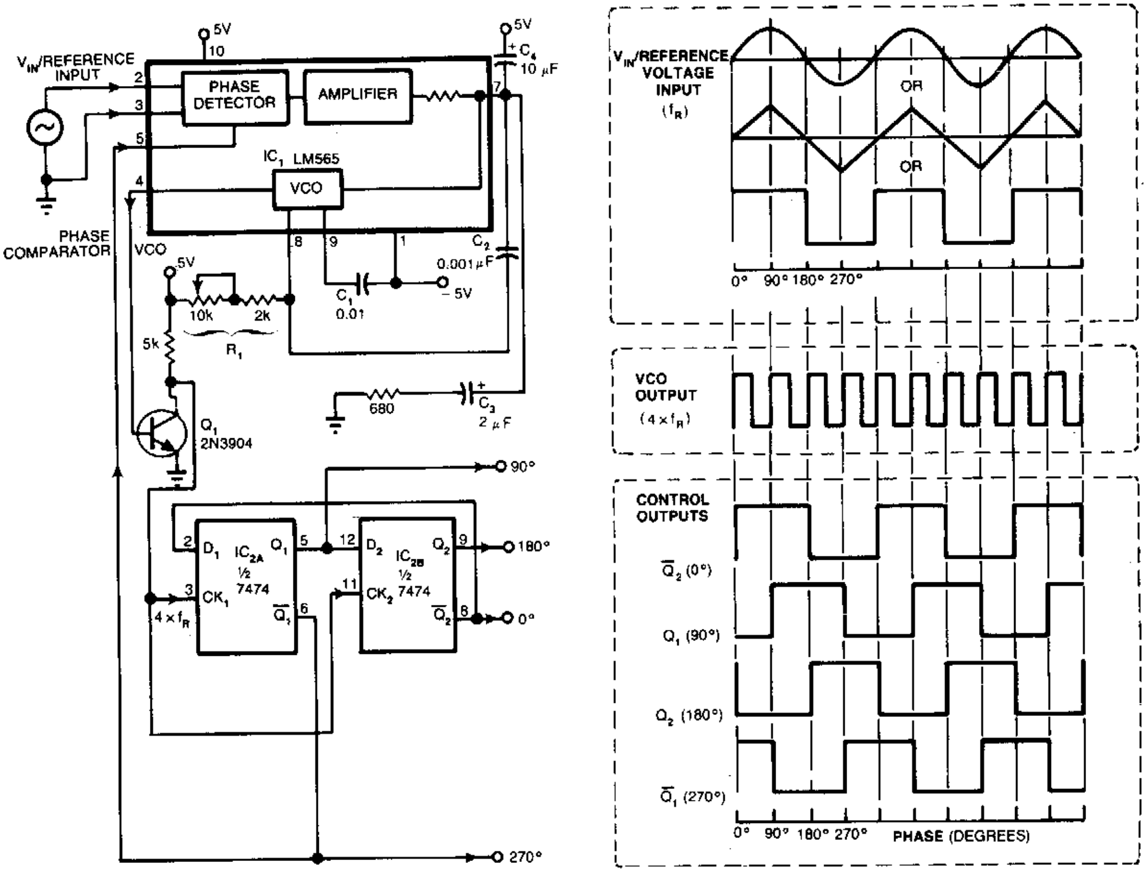


EDN

Fig. 38-2

Using one current source for the charge and discharge path in this circuit ensures identical rise and fall times at the capacitor terminal. A Darlington pair ensures identical biasing of the IC during the charge and discharge cycles. The period of the triangle wave is: T ≈ $0.46VC1/R2$. V_{CC} must be at least 8 V to maintain linearity. At the output at pin 3 of the IC timer, a 50% duty-cycle square wave, frequency tunable by R2 alone, appears.

FOUR-OUTPUT WAVEFORM GENERATOR

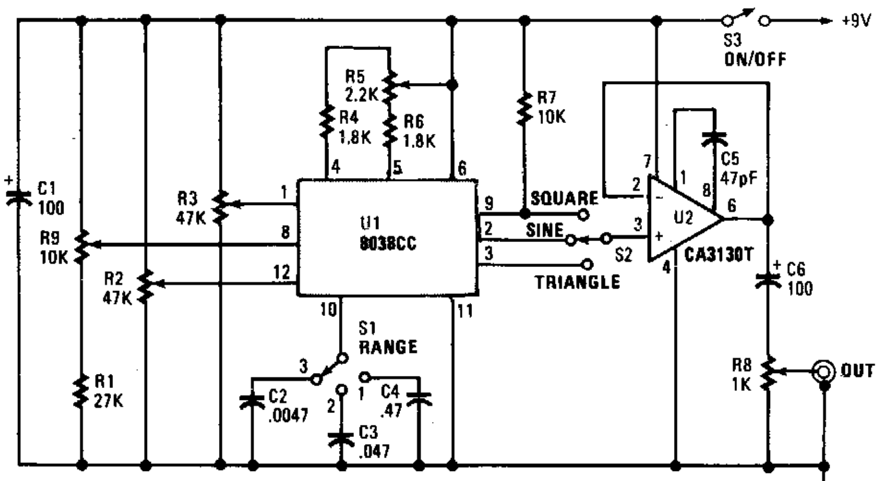


(A) (B)

EDN *Fig. 38-3*

Many applications require control signals that have phase shifts with reference to an input signal. Circuit accepts a sine, square, or triangular wave as an input reference signal and produces square-wave outputs with 0°, 90°, 180°, and 270° phase shifts with respect to the input. Figure 38-3B shows the input and output waveforms. The circuit contains two ICs: an LM565 phase-locked loop and a 7474 dual-D positive edge-triggered flip-flop. R1 and C1 set the free-running frequency of the LM565's VCO. You should adjust R1 so that the frequency is approximately four times that of the input reference signal. The LM565 responds to input signals greater than 10 mV pk-pk; 3 V pk-pk is the chip's maximum allowable input level. Q1 matches the LM565's output to the flip-flops' inputs. The flip-flops' outputs provide the TTL-compatible square-wave signals with 0°, 90°, 180°, and 270° phase shift with reference to the input signal.

FUNCTION GENERATOR

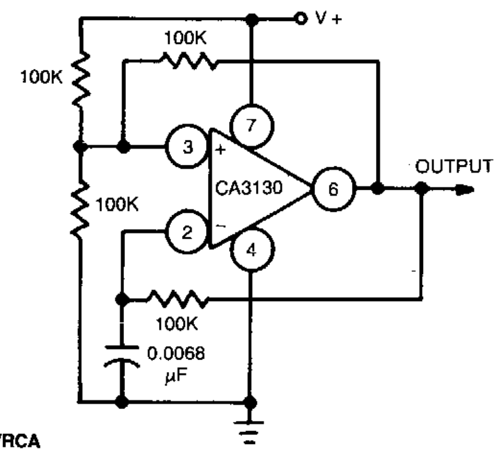


POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

Fig. 38-4

This circuit can output sine, square, and triangular signals of from 15 Hz to 25 kHz in three ranges. The circuit is built around an 8038 function generator that produces the triangular- and square-wave outputs directly from an oscillator. The triangular output is then processed to develop the sine wave. While that method doesn't provide a sufficiently low level of distortion to let you make distortion measurements on audio gear, the degree of purity is high enough for frequency-response tests and a lot of other audio analysis. Three switched capacitors, C2 to C4, set the circuits frequency range via switch S1. Variable resistor R9 and resistor R1 provide the voltage for controlling the charge and discharge rates of the timing capacitor selected. Resistors R4 to R6 control the charge and discharge currents. Resistor R5 can be adjusted to provide a 1.1 mark/space ratio.

CLASSIC OP AMP ASTABLE MULTIVIBRATOR



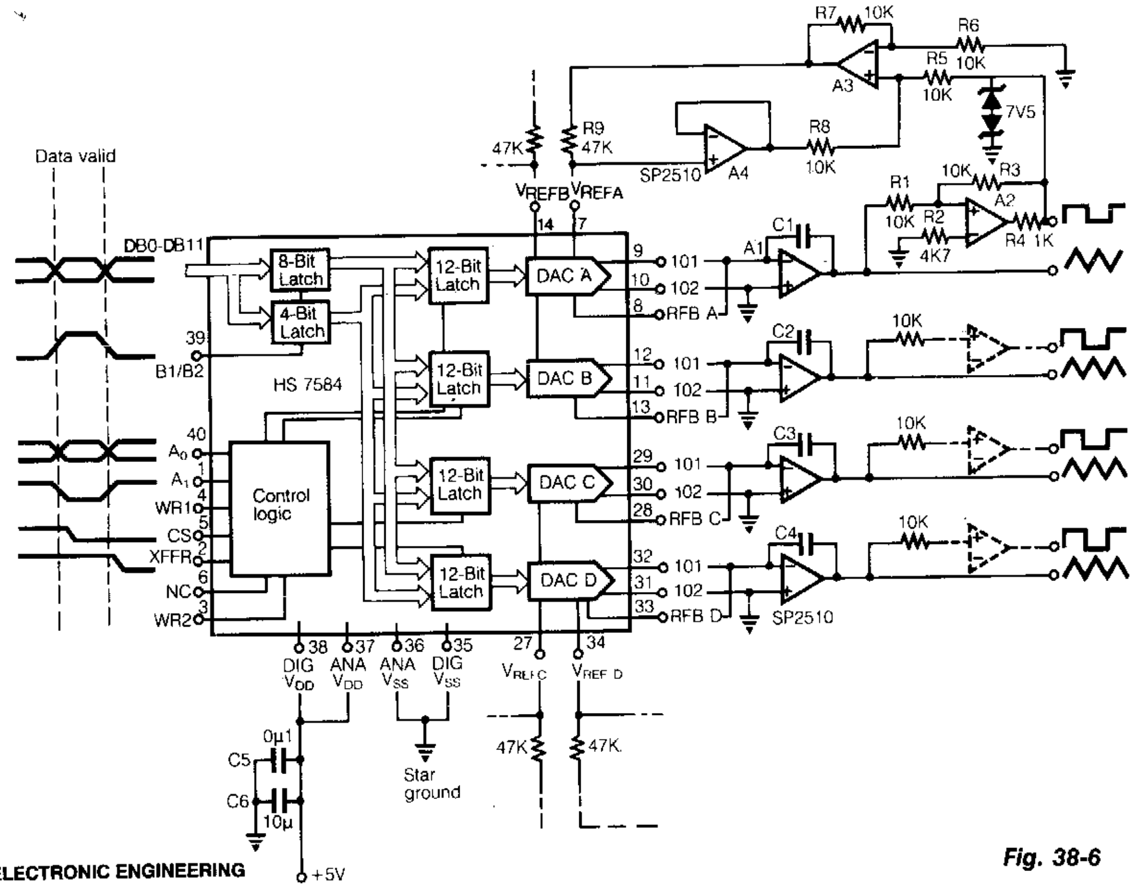
Uses CA3130 BiMOS op amp that operates at a frequency of 1 kHz. With rail-to-rail output swing, frequency is independent of supply voltage, device, and temperature. Only the temperature coefficient of R_T and C_T enters into circuit stability.

GE/RCA

ALL RESISTANCE VALUES ARE IN OHMS

Fig. 38-5

PROGRAMMABLE TRIANGLE-/SQUARE-WAVE GENERATOR



ELECTRONIC ENGINEERING

Fig. 38-6

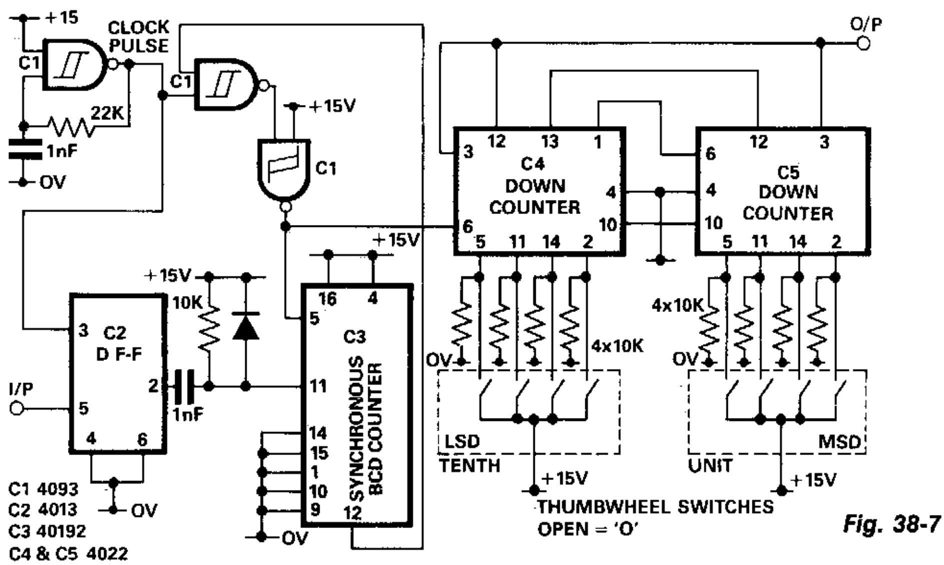
The programmable multiple output generator provides the control signals for data converter ATE. Major performance criteria are simple, interfaces to a number of microprocessor systems, low power consumption, stable output timing relationships combined with a minimum of board space. For schematic simplicity only, one output circuit is shown in full.

The monolithic HS7584 provides four current output DAC's with four quadrant multiplication, individual reference input and a feedback resistor. The digitally controlled integrator's frequency is determined by:

$$f = \frac{\text{Digital Input}}{4RC}$$

C is the value of C1 to C4 and *R* is the resistance of the DAC. With the four DACS on a single chip, the resistance matching is good, which results in stable timing relationships of the generator outputs. The output of the comparator A2 determines whether the constant current source provided by A3 and A4 is positive or negative.

NONINTEGER PROGRAMMABLE PULSE DIVIDER

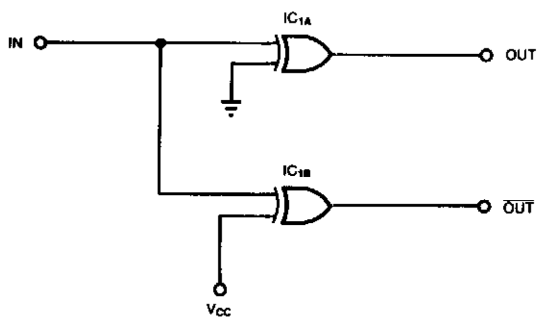


ELECTRONIC ENGINEERING

Fig. 38-7

The purpose of D-type flip-flop IC2 is to synchronize the input signal with the clock pulse. When the clock pulse changes from low to high and the input is high, IC2 output is high. Subsequently, IC3 resets to zero and starts counting up. Until the counter counts to ten, the counter is inhibited. Thus, the number of pulses of the output of IC3 is ten times input pulse. The designed frequency of the clock pulse must be ten times higher than the maximum frequency of the input. IC4 and IC5 are cascaded to form a two decade programmable down counter. Since the number of pulses appearing at the input of the down counter is ten times the input to the divider, the effective range of the divisor for this divider is 0.1 to 9.9.

XOR GATE COMPLEMENTARY SIGNALS GENERATOR

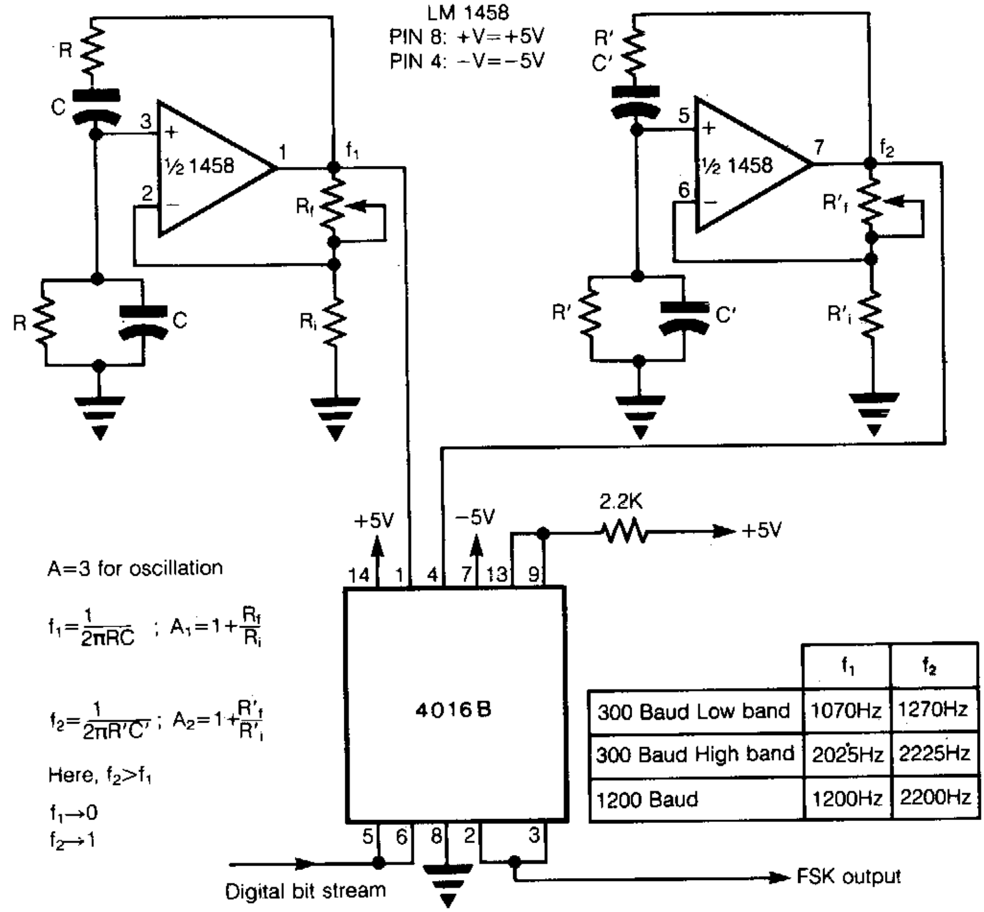


EDN

Fig. 38-8

Some applications, such as driving three-state buffers for data multiplexers or for biphase clocks in high-speed systems, require complementary signals having a small-time skew and nearly simultaneous transitions. Here, XOR gates function as both inverting and noninverting gates. For CMOS systems, practically any type of XOR gate will work. However, the advanced-CMOS logic (ACL) families have the greatest drive capability, the shortest gate delays, and the tightest manufacturing tolerances. For TTL systems, compatible CMOS types such as the ACT or S/AS86 families are preferable. Do not use low-power TTL versions (LS or ALS), because they have large propagation delay differences when one XOR gate is inverting and the other is noninverting.

LOW-COST FSK GENERATOR



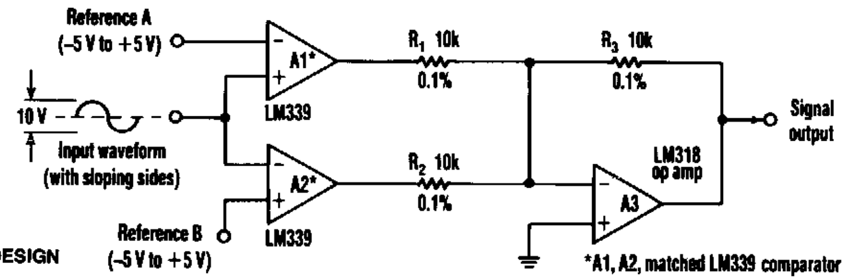
	f_1	f_2
300 Baud Low band	1070Hz	1270Hz
300 Baud High band	2025Hz	2225Hz
1200 Baud	1200Hz	2200Hz

ELECTRONIC ENGINEERING

Fig. 38-9

In FSK, two discrete frequencies are used to represent the binary digits 0 and 1. The heart of the circuit consists of two Wien-bridge oscillators built using a dual op amp LM 1458, for the two frequencies. The two frequencies are enabled corresponding to digital data using two switches in SCL 4016. The control lines of these switches are logically inverted with respect to each other using one of the switches in SCL 4016 as an inverter, so as to enable only one oscillator output at a time. The digital bit stream is used to control the analog switches as shown. Since the switching frequency limit of SCL 4016 is 40 MHz, high-data rates can be easily accommodated. This method comes in handy when expensive FSK generator chips are not readily available; also, the components used in this circuit are easily available off the shelf and are quite cheap.

HARMONICS GENERATOR



ELECTRONIC DESIGN

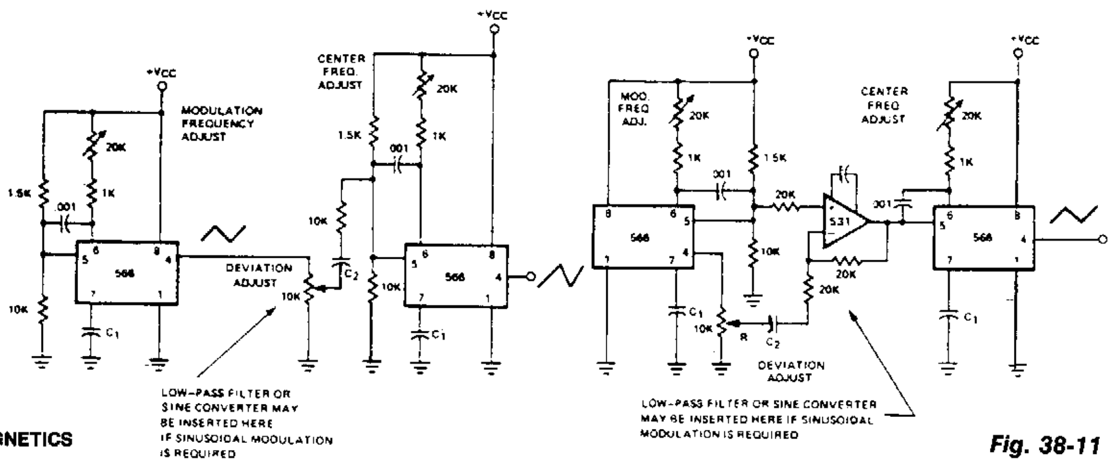
Fig. 38-10

Two comparators and a summing amplifier that generate differential harmonic spectra comprise a simple frequency multiplier. The resulting circuit can extract harmonics from a sine, triangle, sawtooth, or any other sloping-sided waveform.

With a sloped-input waveform, a comparator produces an output pulse width that's proportional to the input amplitude plus a reference voltage. Changing the reference can vary the pulse width from 0 to 100%. As the pulse width changes, the harmonic spectrum changes, but two comparators combined in the adder eliminate harmonics, depending on the duty cycle. For example, a 50% pulse will lack all the even-numbered harmonics. Similarly, a 25% duty-cycle pulse will be missing multiples of the fourth harmonic and deliver the second, sixth, and tenth harmonics. Accordingly, the circuit generates multiples of the input frequency that might not have existed in the input waveform. Adjusting the references can create virtually any harmonic.

Because comparators A1 and A2 supply differential inputs to the added A3, the adder cancels out equal harmonics. Therefore, both A1 and A2 should have identical ac characteristics, and A3 should have good common-mode rejection and a high slew rate. In particular, R1, R2, and R3 should match within 0.1%. Of course, the accuracy of the circuit depends heavily on the amplitude stability of the input.

LOW-FREQUENCY FM GENERATORS



SIGNETICS

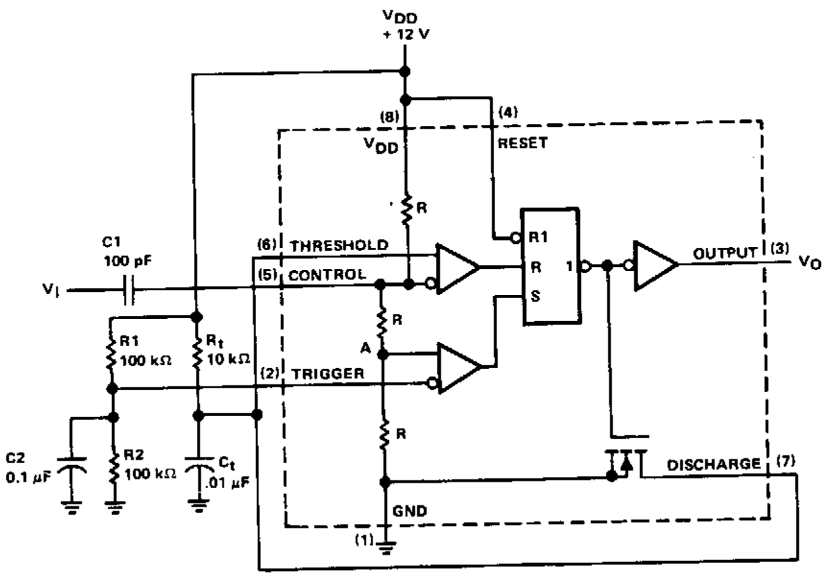
a. Small Frequency Deviations to ±20%

b. Large Frequency Deviations to ± 100%

Fig. 38-11

Here are two FM generators for low frequency, less than 0.5 MHz center frequency, applications. Each uses a 566 function generator as a modulation generator and a second 566 as the carrier generator. Capacitor C1 selects the modulation frequency adjustment range and C1 selects the center frequency. Capacitor C2 is a coupling capacitor which only needs to be large enough to avoid distorting the modulating waveform. If a frequency sweep in only one direction is required, the 566 ramp generators given in this section can be used to drive the center generator.

POSITIVE-TRIGGERED MONOSTABLE



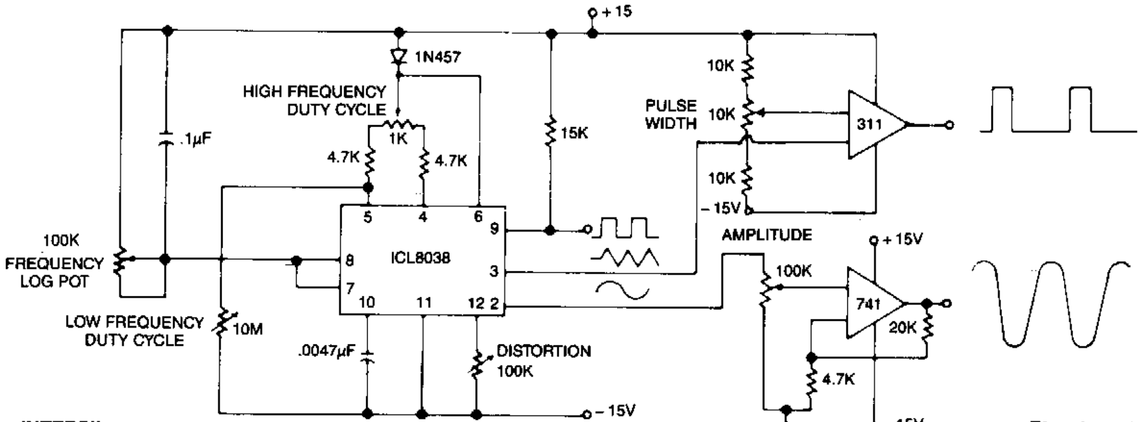
Reprinted by permission of Texas Instruments.

Fig. 38-12

A positive-going trigger pulse can be used to start the timing cycle with the circuit shown. In this design, trigger input pin 2 is biased to 6 V (1/2 V_{DD}) by divider R1 and R2. Control input pin 5 is biased to 8 V (2/3 V_{DD}) by the internal divider circuit. With no trigger voltage applied, point A is at 4 V (1/3 V_{DD}). To turn the timer on, the voltage at point A has to be greater than the 6 V present on pin 2. Positive 5-V trigger pulse V_I applied to the control input pin 5 is ac coupled through capacitor C1, adding the trigger voltage to the 8 V already on pin 5; this results in 13 V with respect to ground. The output pulse width is determined by the values of R_t and C_t.

When voltage at point A is increased to 6.5 V, which is greater than the 6 V on pin 2, the timer cycle is initialized. The output of timer pin 3 increases, turning off discharge transistor pin 7 and allowing C_t to charge through resistor R_t. When capacitor C_t charges to the upper threshold voltage of 8 V (2/3 V_{DD}), the flip-flop is reset and output pin 3 decreases. Capacitor C_t then discharges through the discharge transistor. The timer is not triggered again until another trigger pulse is applied to control input pin 5.

PRECISION AUDIO WAVEFORM GENERATOR

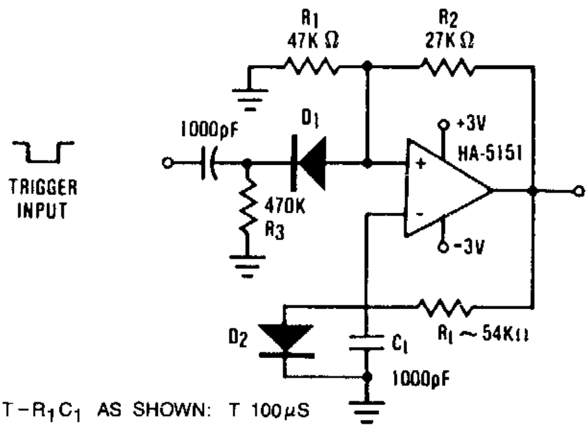


INTERSIL

Fig. 38-13

This circuit generates sinusoidal, square, and triangle waveforms simultaneously. Set the frequency to a particular value or vary it, as shown above. An op amp can be added for extra drive capability and simplified amplitude adjustment. A simple comparator, slicing the triangle waveform, provides continuous duty cycle adjustment at a constant frequency.

MONOSTABLE MULTIVIBRATOR



T - R1C1 AS SHOWN: T 100μS

HARRIS

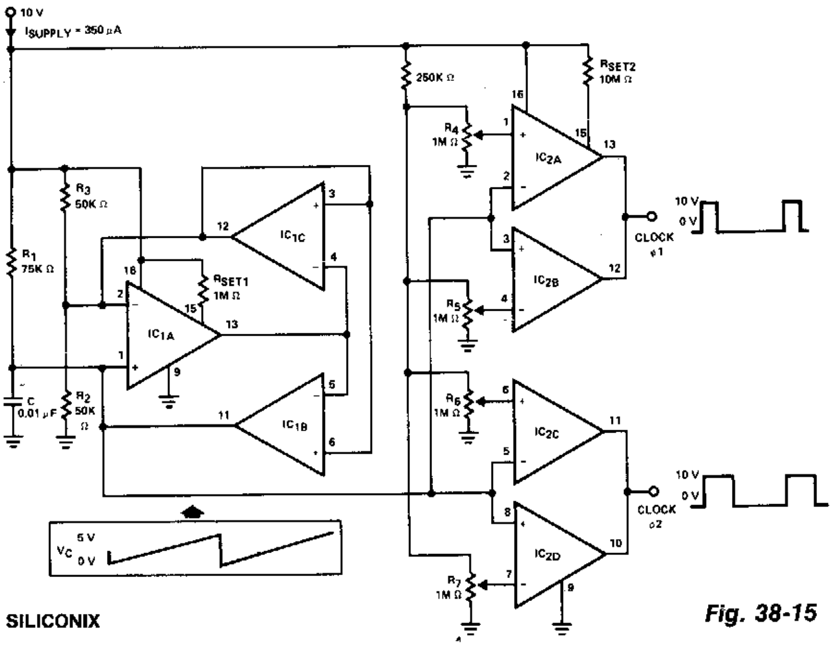
Fig. 38-14

The circuit illustrates the usefulness of the HA-5151 as a battery-powered monostable. In this circuit, the ratio is set to .632, which allows the time constant equation to be reduced to:

$$T = R_t C_t$$

D2 is used to force the output to a defined state by clamping the negative input at +0.6 V. Triggering is set by C1, R3, and D2. An applied trigger pulls the positive input below the clamp voltage, +0.6 V, which causes the output to change state. This state is held because the negative input cannot follow the change because of $R_t \cdot C_t$. This particular circuit has an output pulse width set at approximately 100 μs. Use of potentiometers for R_t and variable capacitors for C_t will allow for a wide variation in T.

VERSATILE 2ϕ PULSE GENERATOR



SILICONIX

Fig. 38-15

FIXED-FREQUENCY GENERATOR

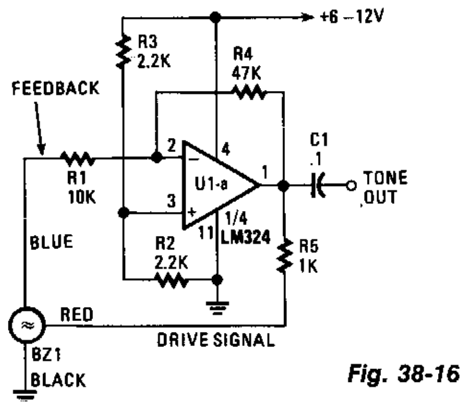


Fig. 38-16

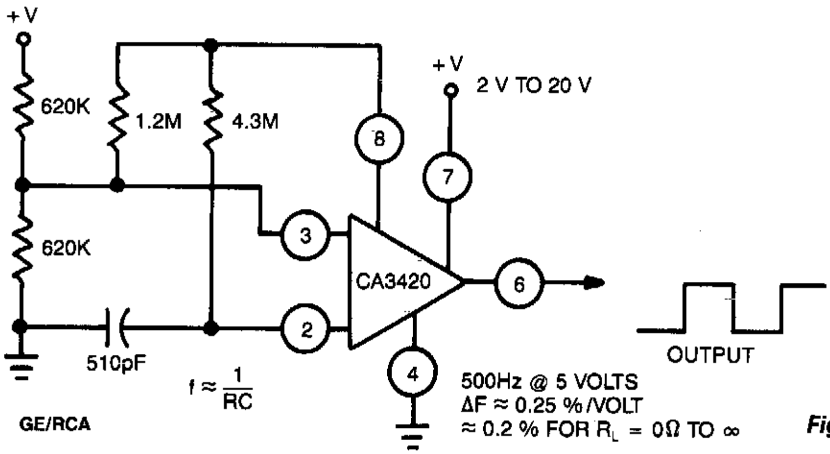
HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

A single op amp, one fourth of an LM324 quad op amp, is configured as a standard inverting amplifier. At power up, a positive voltage is applied to the noninverting input of U1, via R3, forcing its output high. That high output travels along three paths. The first path is the tone output. Along the second path, by way of R5, that high is used as the drive signal for BZ1. In the third path, the high output of U1 is fed back, via R4, to the inverting input of U1. That forces U1's output to go low. And that low, when fed back to the inverting input of U1, causes the op amp output again to a high, and the cycle repeats itself. As configured, U1 provides a voltage gain of 4.7 (gain = R4/R1).

The outer ring of the piezo element is usually connected to the circuit ground. The large inner circle serves as the driven area, and the small elongated section supplies the feedback signal. Resistor R5 sets BZ1's output-volume level. That level can be increased by decreasing R5 for example, to 470 Ω. To decrease the volume, increase R5 to about 2.2 KΩ, or so.

Resistors R2 and R3 set the bias for op amp U1's positive input pin 3 to half of the supply-voltage level. That allows for a maximum voltage swing at U1's output. Although a quad op amp is specified, almost any similar low cost single or dual op amp will work for U1a.

SINGLE-SUPPLY MULTIVIBRATOR

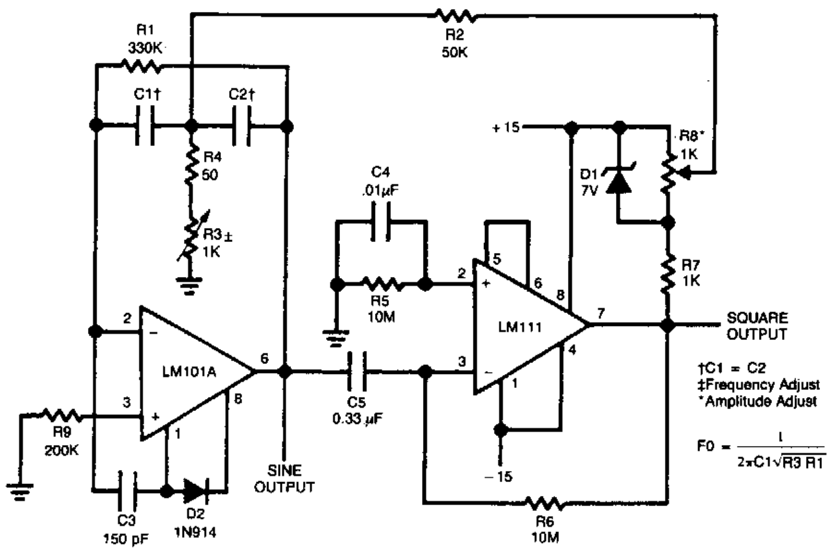


GE/RCA

Fig. 38-17

This multivibrator uses a CA3420 BiMOS op amp to provide improved frequency stability. The output frequency remains essentially independent of supply voltage. Because of the inherent buffering action of pin 6, frequency shift is approximately 0.2% when R_L varies between zero Ω to infinity.

EASILY TUNED SINE-WAVE/SQUARE-WAVE OSCILLATOR



NATIONAL SEMICONDUCTOR CORP.

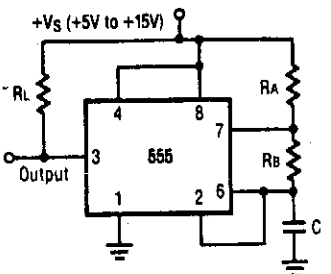
Fig. 38-18

EASILY TUNED SINE-WAVE / SQUARE-WAVE OSCILLATOR (Cont.)

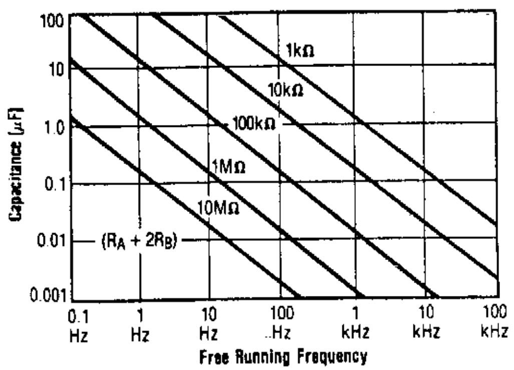
The circuit will provide both a sine- and square-wave output for frequencies from below 20 Hz to above 20 kHz. The frequency of oscillation is easily tuned by varying a single resistor. This is a considerable advantage over Wien-Bridge circuits where two elements must be tuned simultaneously to change frequency. Also, the output amplitude is relatively stable when the frequency is changed. An amp is used as a tuned circuit, driven by square wave from a voltage comparator. The frequency is controlled by R1, R2, C1, C2, and R3, with R3 used for tuning. Tuning the filter does not affect its gain or bandwidth, so the output amplitude does not change with frequency.

A comparator is fed with the sine-wave output to obtain a square wave. The square wave is then fed back to the input of the tuned circuit to cause oscillation. Zener diode, D1, stabilizes the amplitude of the square wave fed back to the filter input. Starting is insured by R6 and C5 which provide dc negative feedback around the comparator. This keeps the comparator in the active region. Distortion ranges between 0.75% and 2% depending on the setting of R3. Although greater tuning range can be accomplished by increasing the size of R3 beyond 1 KΩ, distortion becomes excessive. Decreasing R3 lower than 50 Ω can make the filter oscillate by itself.

ASTABLE



Free Running Frequency vs. R_A, R_B and C

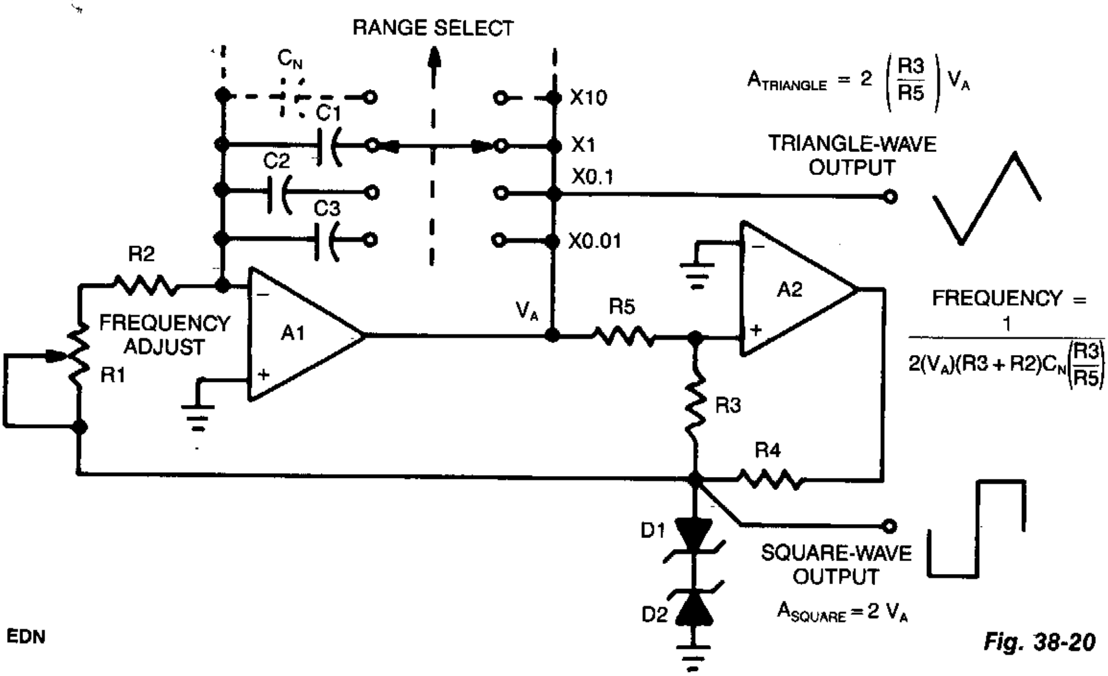


Reprinted with permission from Raytheon Co., Semiconductor Division.

This astable will trigger itself and run free as a multivibrator. The external capacitor charges through R_A and R_B and discharges through R_B only. Thus, the duty cycle is set by the ratio of these two resistors, and the capacitor charges and discharges between 1/3 V_S and 2/3 V_S. The charge and discharge times, and therefore frequency, are independent of supply voltage. The free-running frequency versus R_A, R_B and C is shown in the graph.

Fig. 38-19

TWO-FUNCTION SIGNAL GENERATOR

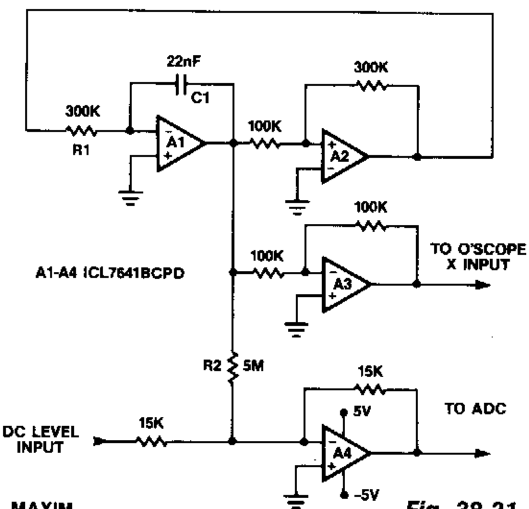


EDN

Fig. 38-20

You can continuously vary the frequencies of the triangle and square waves produced by this circuit over a full decade. If R5 = R3, the amplitude of the two waveforms will be equal ($A_{SQUARE} = A_{TRIANGLE}$).

TRIANGLE GENERATOR

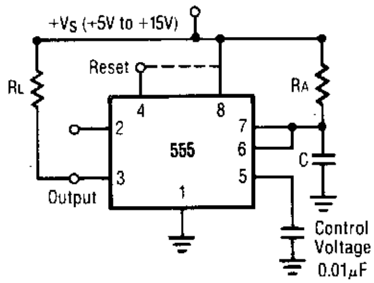


MAXIM

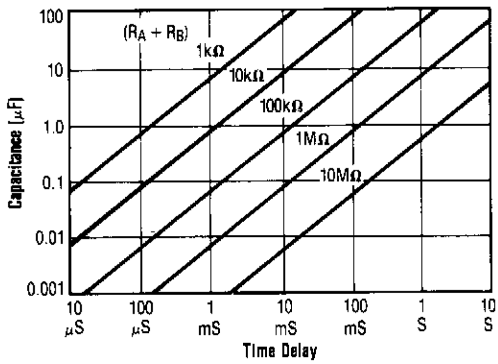
Fig. 38-21

This circuit generates a symmetrical, 10-mV pk-pk triangle waveform which is summed with a dc level and connected to the a/d analog input for noise/DNL testing. The dc level input offsets the triangle waveform over the input range of the ADC. The 10-mV amplitude amounts to an 8 LSB span for a 12-bit, 5-V, full-scale ADC.

MONOSTABLE OPERATION



Time Delay vs. R_A, R_B and C



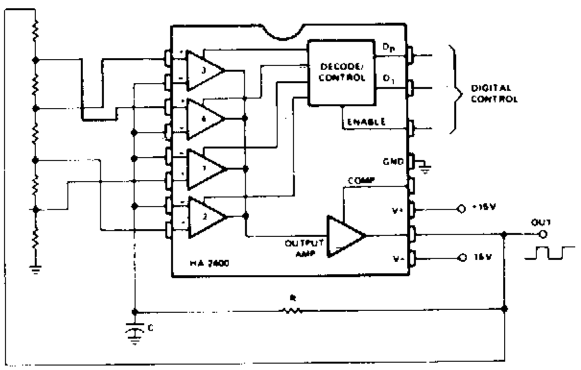
Reprinted with permission from Raytheon Co., Semiconductor Division.

Fig. 38-22

In this mode, the timer functions as a one shot. The external capacitor is initially held discharged by a transistor internal to the timer. Applying a negative trigger pulse to pin 2 sets the flip-flop, driving the output high, and releasing the short circuit across the external capacitor. The voltage across the capacitor increases with the time constant $r = R_AC$ to $^2/_3\ V_S$, where the comparator resets the flip-flop and discharges the external capacitor. The output is now in the low state.

Circuit triggering takes place when the negative-going trigger pulse reaches $^1/_3\ V_S$; the circuit stays in the output high state until the set time elapses. The time the output remains in the high state is 1.1 R_AC and can be determined by the graph. A negative pulse applied to pin 4 (reset) during the timing cycle will discharge the external capacitor and start the cycle over again beginning on the positive-going edge of the reset pulse. If reset function is not used, pin 4 should be connected to V_S to avoid false resetting.

PROGRAMMABLE-FREQUENCY, FREE-RUNNING MULTIVIBRATOR

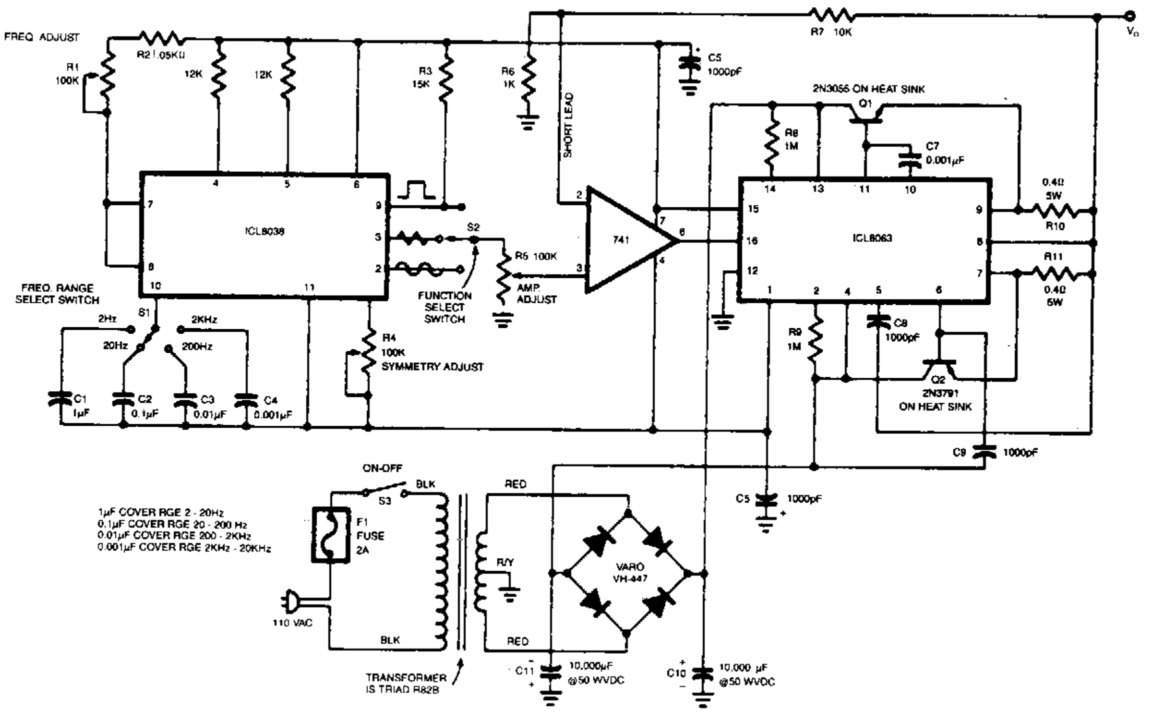


HARRIS

Fig. 38-23

This is the simplest of any programmable oscillator circuit, since only one stable timing capacitor is required. The output square wave is about 25 V pk-pk, and has rise and fall times of about 0.5 μs. If a programmable attenuator circuit is placed between the output and the divider network, 16 frequencies can be produced with two HA-200's and still only one timing capacitor.

FUNCTION GENERATOR

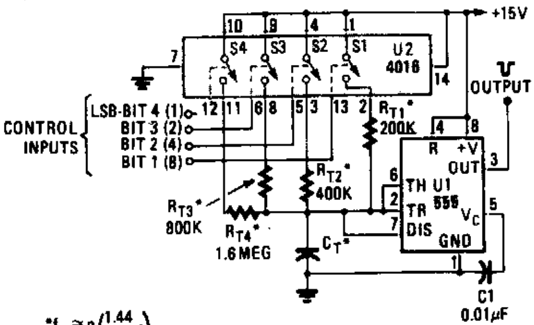


INTERSIL *Fig. 38-24*

This generator will supply sine, triangular, and square waves from 2 Hz to 20 kHz. This complete test instrument can be plugged into a standard 110 Vac line for power. V_{OUT} will be up to ±25 V (50 V pk-pk across loads as small as 10 Ω (about 2.5 A maximum output current).

Capacitor working voltages should be greater than 50 Vdc and all resistors should be 1/2 W, unless otherwise indicated. The interconnecting leads from the 741 pins 2 and 3 to their respective resistors should be kept short, less than 2 inches if possible; longer leads might result in oscillation. Full output swing is possible to about 5 kHz; after that the output begins to taper off because of the slew rate of the 741, until at 20 kHz the output swing will be about 20 V_{pp} ± 10 V. This problem can be remedied by simply using an op amp with a higher slew rate, such as the LF356.

PROGRAMMABLE-FREQUENCY ASTABLE



$$*f_o \cong n\left(\frac{1.44}{R_T C_T}\right)$$

WHERE,

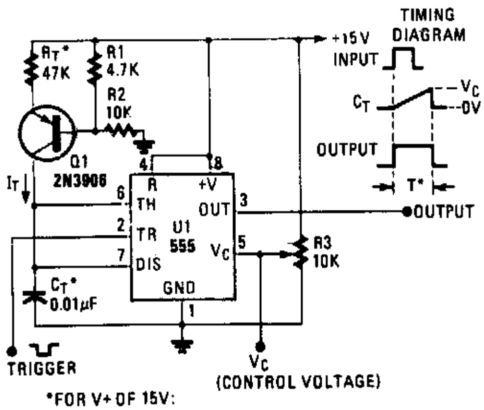
n is DIGITAL INPUT WORD: $1 \leq n \leq 15$

(AS SHOWN, WITH BASE R_T OF 1.6 MEG, 100 Hz $\leq f_o$ 1500 Hz).

POPULAR ELECTRONICS

Fig. 38-25

LINEAR-RAMP MONOSTABLE



*FOR V+ OF 15V:

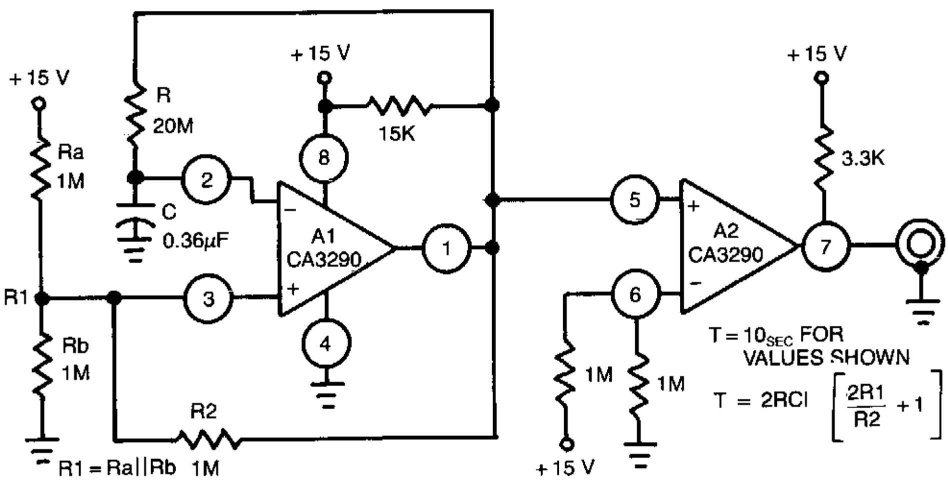
$$T = \frac{V_C C_T}{I_T},\ I_T \cong \frac{4.2}{R_T} \qquad T \cong 0.24 V_C R_T C_T$$

(AS SHOWN, $T_{MAX} \cong 1$ MS WITH V_C = 10V.)

HANDS-ON ELECTRONICS

Fig. 38-26

LOW-FREQUENCY MULTIVIBRATOR



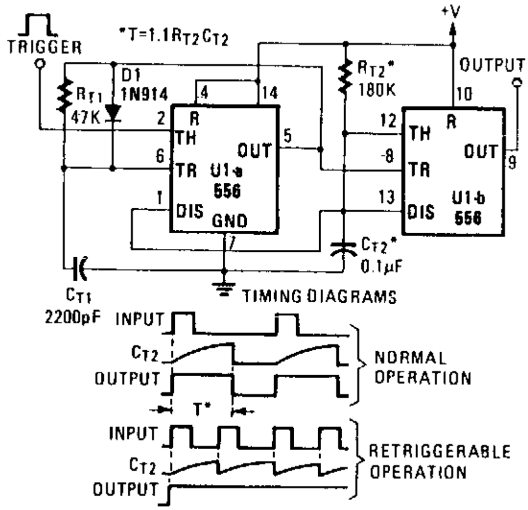
ALL RESISTANCE VALUES ARE IN OHMS

GE/RCA

Fig. 38-27

This circuit uses half the CA3290 BiMOS dual voltage comparator as conventional multivibrator. The second half maintains frequency against effects of output loading. Large values of timing resistor, R1, assure long time delays with low-leakage capacitors.

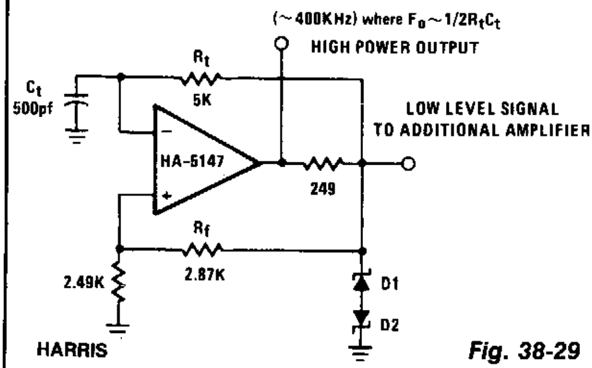
RETRIGGERABLE ONE SHOT



HANDS-ON ELECTRONICS

Fig. 38-28

ASTABLE MULTIVIBRATOR

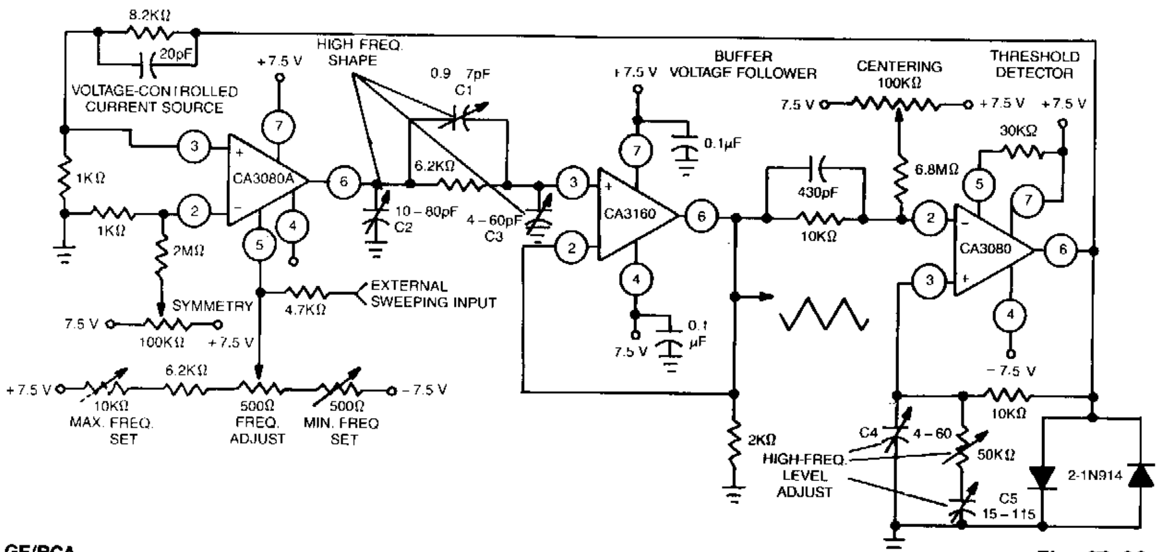


HARRIS

Fig. 38-29

The power bandwidth of the HA-5147 extends the circuit's frequency range to approximately 500 kHz. R_t can be made adjustable to vary the frequency if desired. Any timing errors because of V_{OS} or I_{bias} have been minimized by the precision characteristics of the HA-5147. D1 and D2, if used, should be matched to prevent additional timing errors. These clamping diodes can be omitted by tying R_t and positive feedback resistor R_f directly to the output.

SINGLE-CONTROL FUNCTION GENERATOR

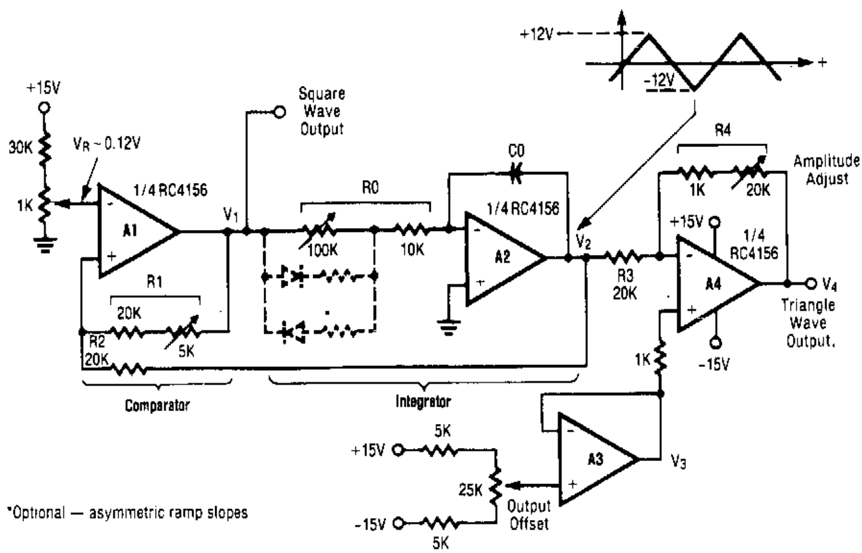


GE/RCA

Fig. 38-30

This function generator, with an adjustment range in excess of 1,000,000 to 1, uses a CA3160 BiMOS op amp as a voltage follower, a CA3080 OTA as a high-speed comparator, and a CA3080 as a programmable-current source. Three variable capacitors, C1, C2, and C3 shape the triangular signal between 500 kH and 1 MHz. Capacitors C4 and C5, and the trimmer potentiometer in series with C5, maintain essentially constant (±10%) amplitude to 1 MHz.

TRIANGLE-/SQUARE-WAVE GENERATOR

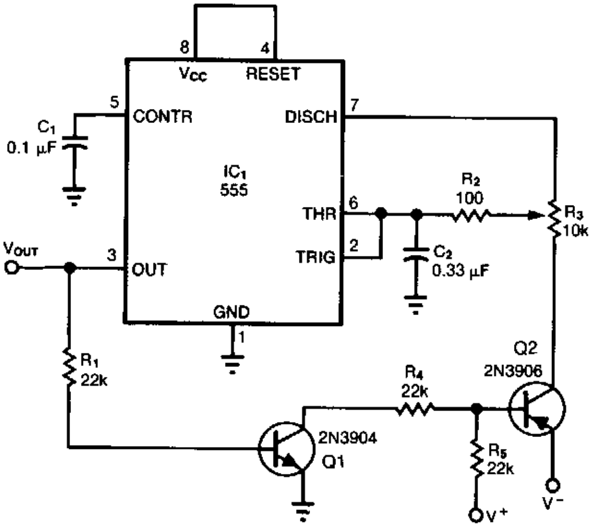


Reprinted with permission from Raytheon Co., Semiconductor Division.

Fig. 38-31

This circuit uses a positive-feedback loop closed around a combined comparator and integrator. When power is applied, the output of the comparator will switch to one of two states, to the maximum positive or maximum negative voltage. This applies a peak input signal to the integrator, and the integrator output will ramp either down or up, opposite of the input signal. When the integrator output, which is connected to the comparator input, reaches a threshold set by R1 and R2, the comparator will switch to the opposite polarity. This cycle will repeat endlessly, the integrator charging positive then negative, and the comparator switching in a square-wave fashion.

VARIABLE DUTY CYCLE TIMER



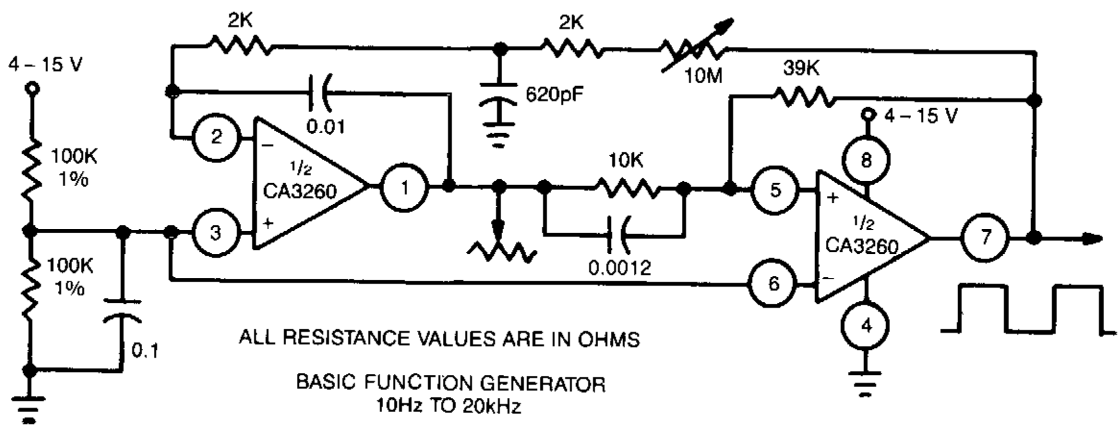
EDN

Fig. 38-32

When configured as a free-running multivibrator, a 555 timer provides no more than a 50% duty cycle. By adding two transistors, however, you can obtain a variable 5 to 95% duty cycle without changing the sum of the on and off times. When V_{OUT} decreases, Q1 is on and Q2 is off, disconnecting V+ while timing capacitor C2 discharges into pin 7 of the timer. When V_{OUT} increases, Q2 reconnects V+ for recharging C2.

Adjusting linear trimming potentiometer R3 to increase the charging resistance increases the on time, but decreases the off time by the same amount by lowering the discharge resistance (the converse is also true). As a result, the sum of the on and off times remains constant. R2 protects Q2 and the timer against high-charge/discharge currents.

BASIC FUNCTION GENERATOR

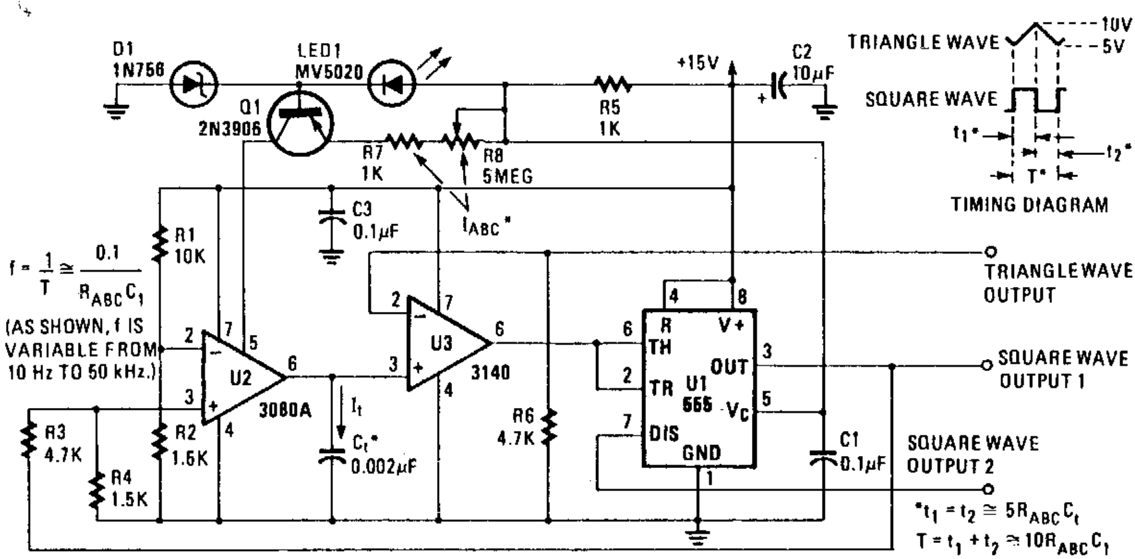


GE/RCA

Fig. 38-33

This function generator uses a CA3260 BiMOS op amp to perform both the integrator and switching functions. A 620-pF capacitor and 2-KΩ resistor shape feedback square wave to reduce spikes. Full audio spectrum, 10 Hz to 20 kHz, is covered with a single 10Ω potentiometer. Requires 9-V battery.

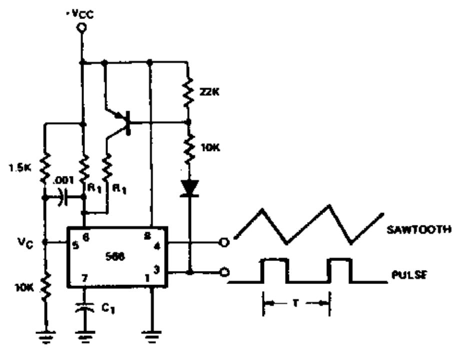
WIDE-RANGE TUNABLE FUNCTION GENERATOR



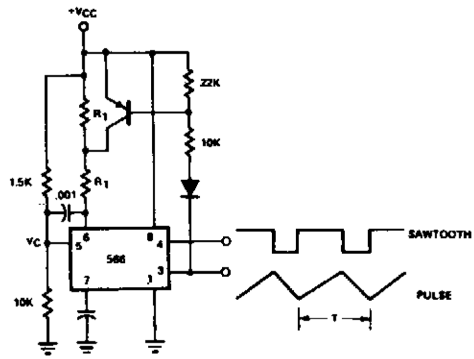
HANDS-ON ELECTRONICS

Fig. 38-34

SAWTOOTH AND PULSE GENERATOR



a. Positive Sawtooth



b. Negative Sawtooth

SIGNETICS

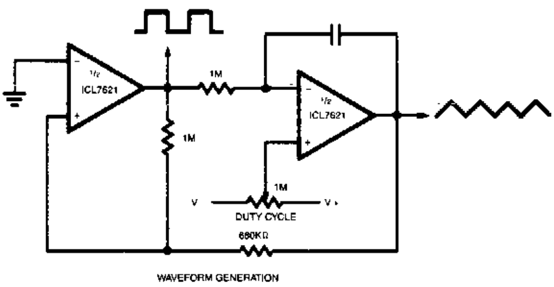
Fig. 38-35

The pin 3 output of the 566 can be used to provide different charge and discharge currents for C1 so that a sawtooth output is available at pin 4 and a pulse at pin 3. The pnp transistor should be well saturated to preserve good temperature stability. The charge and discharge times can be estimated by using the formula:

$$T = \frac{R_T C_1 V_{CC}}{5(V_{CC} - V_C)}$$

where R_T is the combined resistance between pin 6 and V_{CC} for the interval considered.

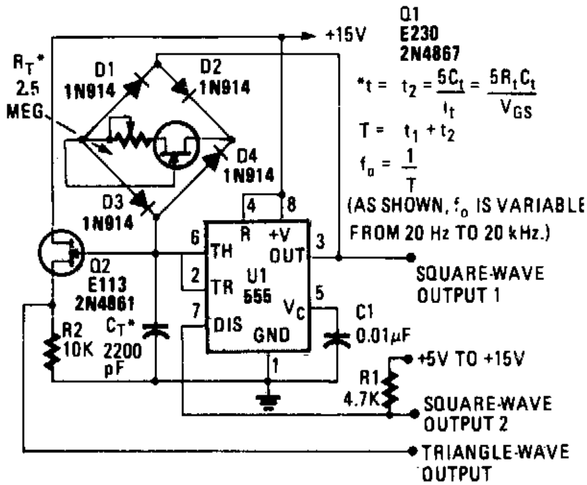
PRECISE TRIANGLE-/ SQUARE-WAVE GENERATOR



INTERSIL *Fig. 38-36*

Since the output range swings exactly from rail to rail, frequency and duty cycle are virtually independent of power supply variations.

WIDE-RANGE TRIANGLE-/ SQUARE-WAVE GENERATOR



HANDS-ON ELECTRONICS *Fig. 38-37*

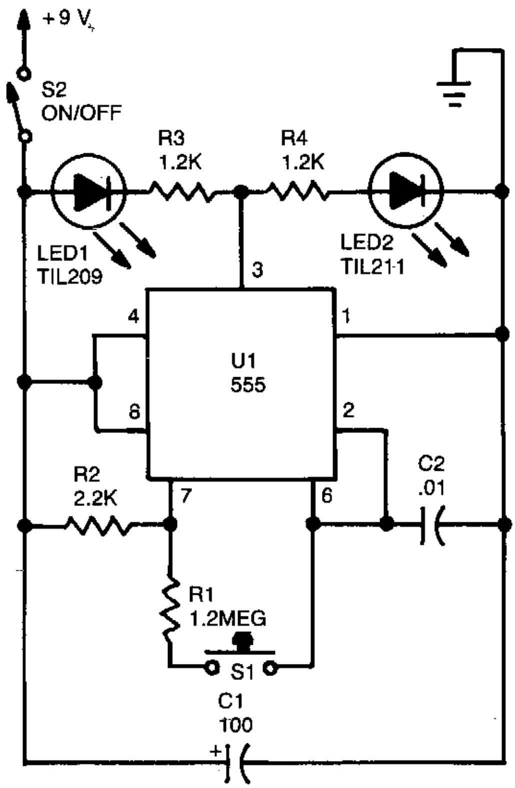
39

Games

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Coin Flipper
Who's First
Electronic Dice

COIN FLIPPER

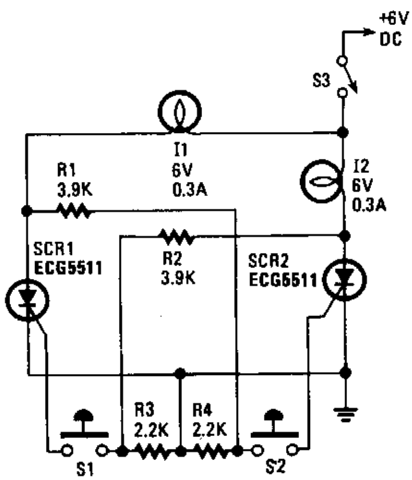


POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

The circuit is basically a 555 astable circuit that divides two LEDs, LED1 and LED2. LED2 is switched on when the output of U1 is high, and LED1 is activated when its output is low. When U1 oscillates, LED1 and LED2 switch on alternately as the output of U1 switches from state to state. Resistor R1's value is high in comparison to R2, so the waveform at the output is a square wave with a mark/space ratio of nearly one-to-one. When you release S1, you break the circuit and U1 latches whatever the output state happens to be at the time.

Fig. 39-1

WHO'S FIRST

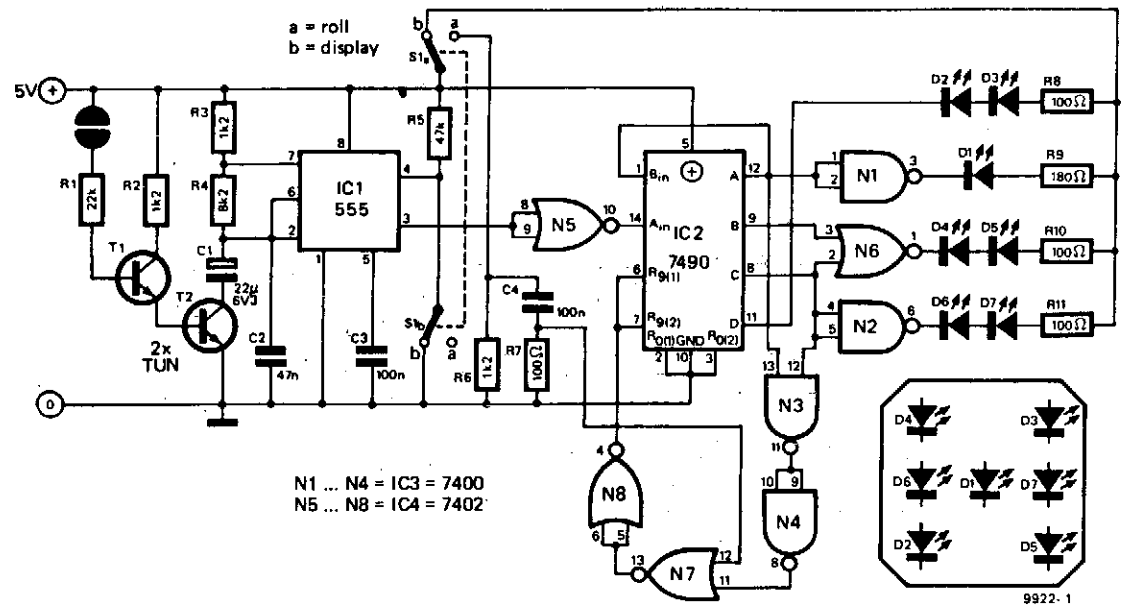


POPULAR ELECTRONICS

Fig. 39-2

Using two SCRs, this control circuit is designed to lock out the other SCR when one has been triggered, so only one lamp will light. Indicator lamp I1 is controlled by SCR1. The operator simply presses switch S1. Lamp I2 is similarly controlled by S2 and SCR2. With both switches open, neither lamp is lit. The result is insufficient gate current to trigger SCR2 into conduction, so lamp I2 does not light. If S2 is pressed first, the reverse situation occurs. Once one of the SCRs is activated, it is necessary to open S3 to turn the light off.

ELECTRONIC DICE



ELEKTOR

Fig. 39-3

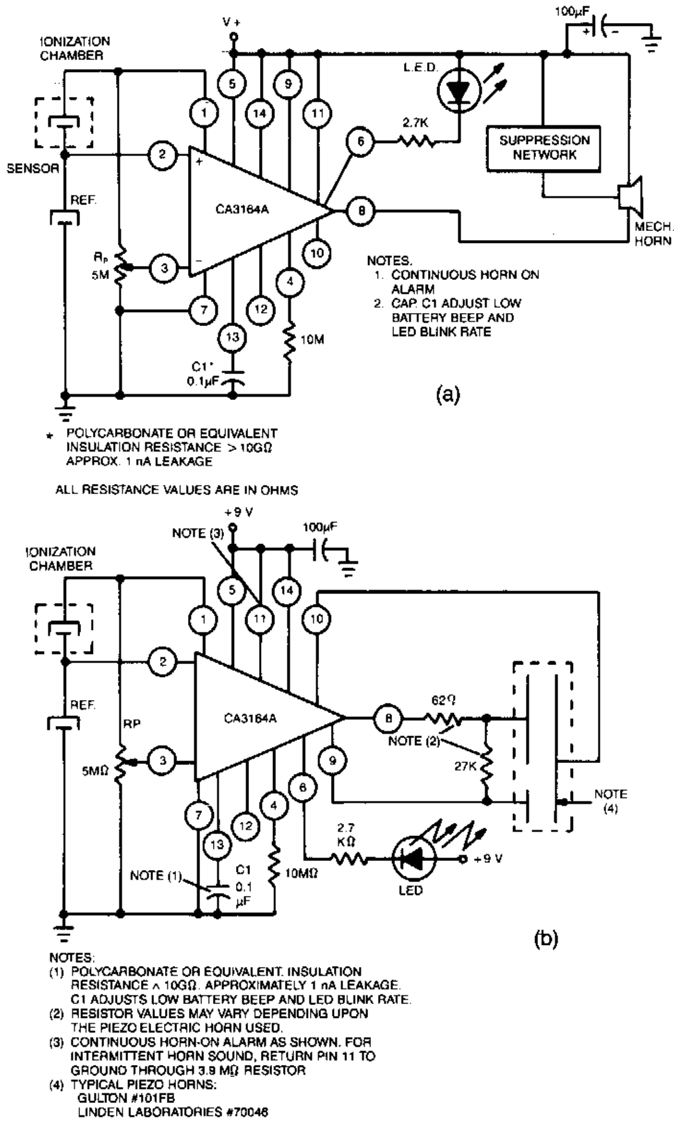
The basic die circuit is given. A 555 timer, IC1, is connected as an astable multivibrator. This feeds clock pulses to divide-by-six counter IC2 the outputs of which are decoded by gates N1 to N6 to drive an array of LEDs in the familiar die pattern. When switch S1 is in position b, the reset input of IC1 decreases and the oscillator is inhibited. Power is fed to the LEDs via S1b so that the display is activated. When the die is *rolled* by switching S1 to position a, the display is blanked. C4 is connected to positive supply via S1a, producing a short pulse which resets IC2 via N7 and N8. The reset input of IC1 is pulled high via R5, so the multivibrator begins to oscillate and feeds clock pulses to IC2 via N5. When S1 is switched back to position a, the multivibrator is again inhibited. Then, the counter stops and power is applied to the LEDs which display the value of the *throw*.

40

Gas and Smoke Detectors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

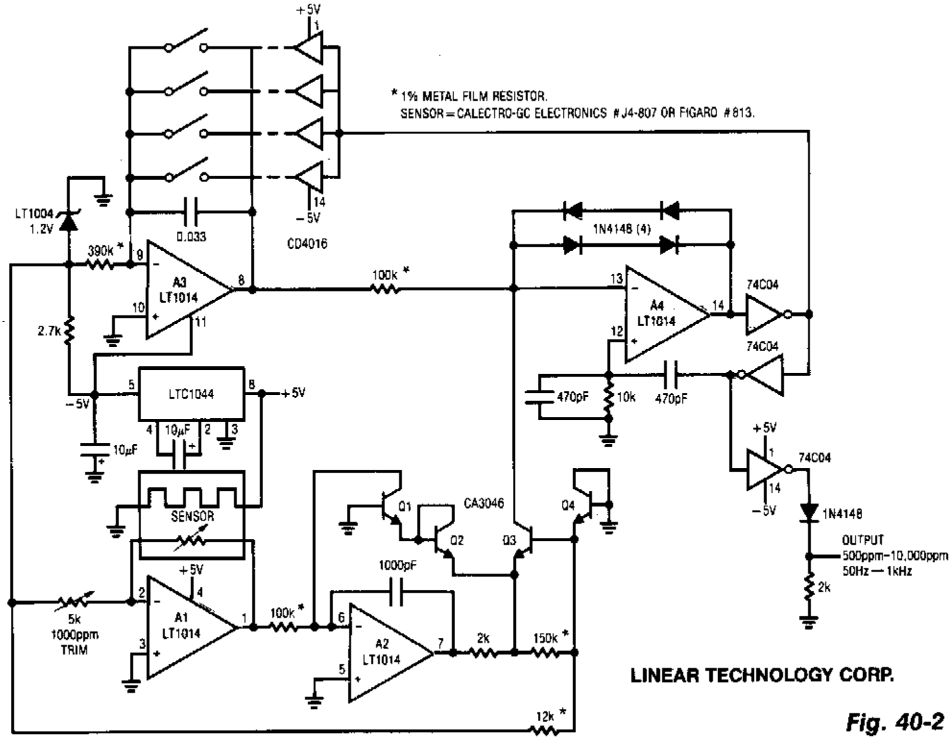
SMOKE DETECTOR



GE/RCA ***Fig. 40-1***

Use CA3164A BiMOS detector/alarm system. For operation as smoke detector with electromechanical horn (Fig. 40-1a), the output of driver at terminal 8 is used. Large npn transistor Q3, with an active pull-up and transistor Q2 provide over 300 mA of drive current. For operation as a smoke detector with a piezoelectric horn (Fig. 40-1b), the circuit requires output from a second inverting amplifier at terminal 10, as well as the output from terminal 8.

METHANE CONCENTRATION DETECTOR WITH LINEARIZED OUTPUT



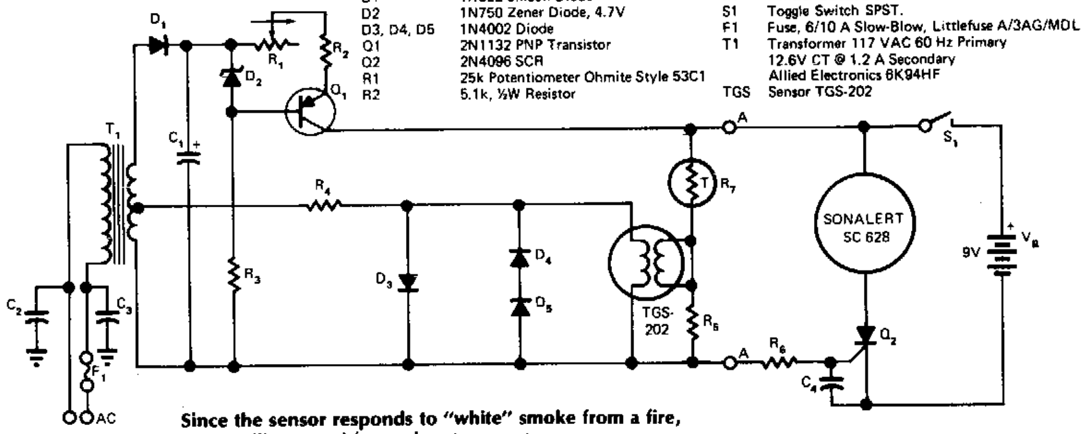
LINEAR TECHNOLOGY CORP.

Fig. 40-2

SMOKE/GAS/VAPOR DETECTOR

PARTS LIST

V_B	9 Volt Transistor Battery	R3	4.7, ½W Resistor
C1	47µF, 35 V	R4	5 Ohms, 12 Watt
C2, C3	100 pf, 600V	R5	68k ½W Resistor
C4	0.01µF, 50V	R6	1.3k ½W Resistor
D1	1N662 Silicon Diode	R7	Fenwal GB32J2 or Veco 35D6
D2	1N750 Zener Diode, 4.7V	S1	Toggle Switch SPST.
D3, D4, D5	1N4002 Diode	F1	Fuse, 6/10 A Slow-Blow, Littlefuse A/3AG/MDL
Q1	2N1132 PNP Transistor	T1	Transformer 117 VAC 60 Hz Primary 12.6V CT @ 1.2 A Secondary Allied Electronics 6K94HF
Q2	2N4096 SCR		
R1	25k Potentiometer Ohmite Style 53C1		
R2	5.1k, ½W Resistor	TGS	Sensor TGS-202



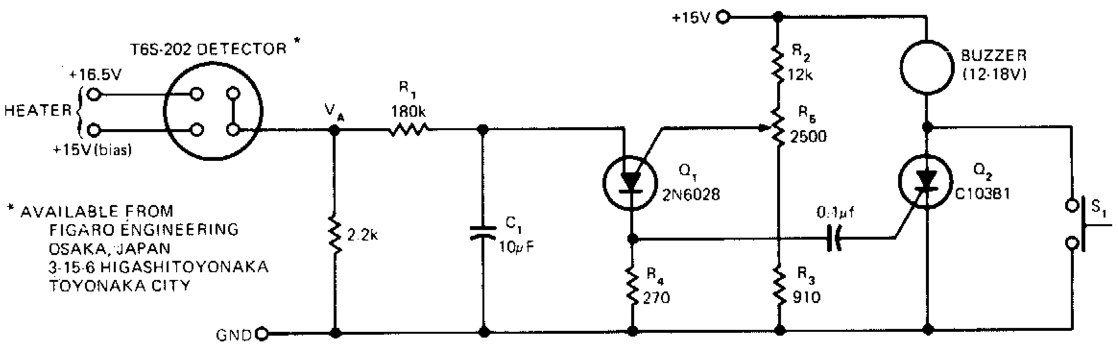
Since the sensor responds to "white" smoke from a fire, alarm will respond faster than temperature sensors.

EDN

Fig. 40-3

Transformer T1 supplies power to the heater of the sensor. Since the sensor is fairly sensitive to heater voltage, diodes D3, D4, and D5 regulate the heater voltage. T1, together with D1 and C2, forms a dc power supply, whose current is regulated by Q1 and adjusted by R1. The constant current from Q1 feeds a variable resistance, consisting of thermistor R7 and the parallel combination of R5 and the sensor resistance. When a hazard causes the voltage at A-A to drop, the net voltage at the SCR gate turns positive, triggering the SCR on and operating the alarm. The alarm draws a small amount of current, so the battery will last a long time. Switch S1 turns off the alarm and resets the SCR.

GAS/SMOKE DETECTOR

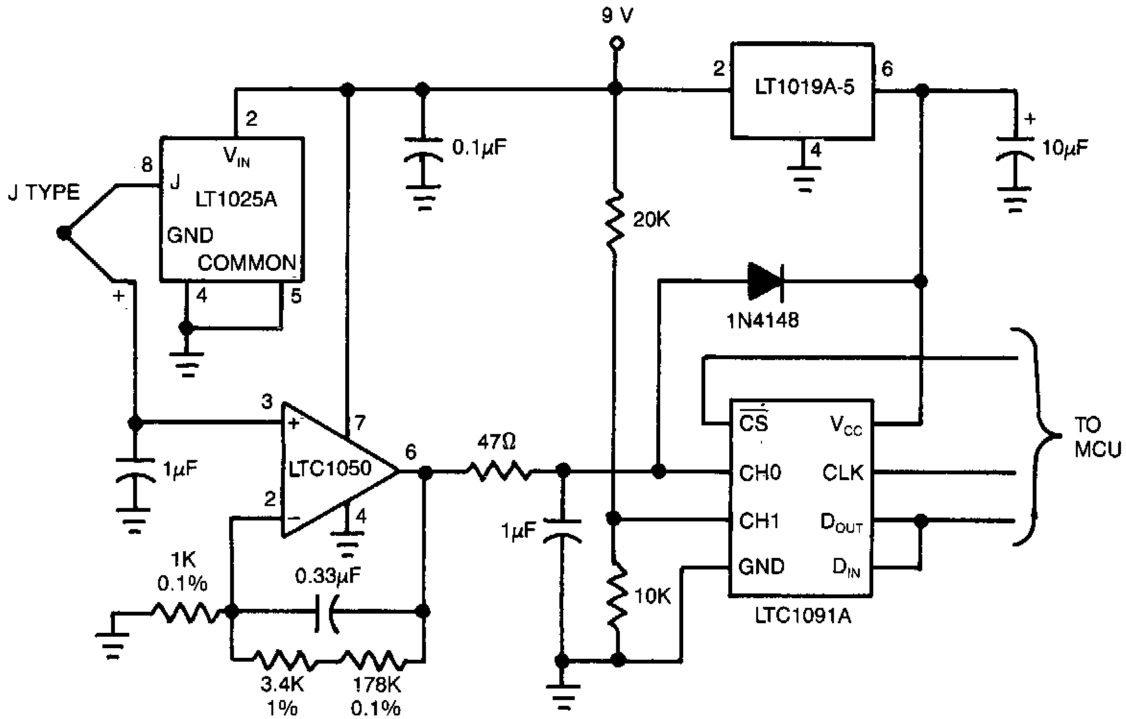


EDN

Fig. 40-4

The sensor is based on the selective absorption of hydrocarbons by an n-type metal-oxide surface. The heater in the device serves to burn off the hydrocarbons once smoke or gas is no longer present in the immediate area; hence, the device is reuseable. When initially turned on, a 15 minute warm-up period is required to reach equilibrium ($V_A \cong 0.6$ V) in a hydrocarbon-free environment. When gas or smoke is introduced near the sensor, V_A will quickly rise (rate and final equilibrium depend on the type of gas and concentration) and trigger Q1, a programmable unijunction transistor. The voltage pulse generated across R4 triggers Q2, sounding the buzzer until S1 resets the unit. R1 and C1 give a time delay to prevent small transient waves of smoke, such as from a cigarette, from triggering the alarm. Triggering threshold is set by R5, R2, and R3; with the components shown, between 50 and 200 ppm of hydrocarbons can be easily detected. Since it is somewhat sensitive to heater voltage, a regulated supply should be used. Power requirements are 1.5 V at 500 mA for the heater and 15 V at 30 mA, depending on type of buzzer, for the bias supply.

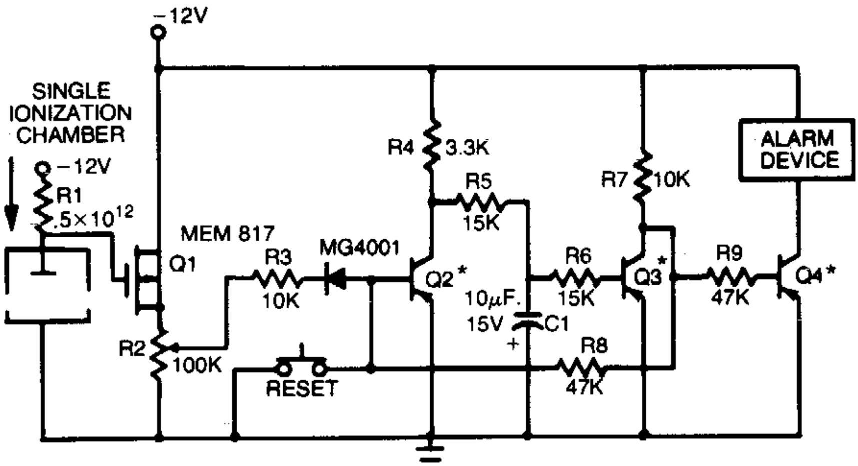
FURNACE EXHAUST GAS TEMPERATURE MONITOR WITH LOW SUPPLY DETECTION



LINEAR TECHNOLOGY DESIGN NOTES *Fig. 40-5*

This circuit can be used to measure exhaust gas temperature in a furnace. The 10-bit LTC1091A gives 0.5°C resolution over a 0°C to 500°C range. The LTC1050 amplifies and filters the thermocouple signal, the LT1025A provides cold junction compensation and the LT1019A provides an accurate reference. The J-type thermocouple characteristic is linearized digitally inside the MCU. Linear interpolation between known temperature points spaced 30°C apart introduces less than 0.1°C error. The 20-K/10 KΩ divider on CH1 of the LTC1091 provides low supply voltage detection. Remote location is easy, with data transferred from the MCU to the LTC1091 via the three-wire serial part.

SMOKE DETECTOR



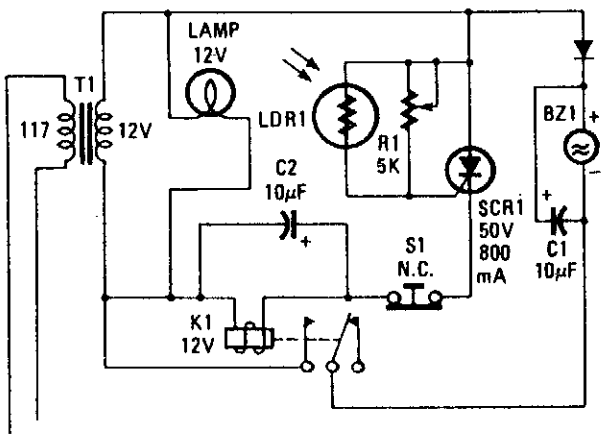
GENERAL INSTRUMENT MICROELECTRONICS

*2N6076 or equivalent

Fig. 40-6

This smoke detector uses a MEM 817 p-channel enhancement mode MOSFET as its buffer amplifier. Operation of the sensor is based on a decrease in the current when smoke enters the chamber, thereby causing a negative voltage excursion at the gate of the buffer MOSFET. Quiescent voltage values at the output of the chamber vary from about −4 V to −6 V, and detection of smoke will result in an excursion of about −4 V. The MOSFET is connected as a source follower.

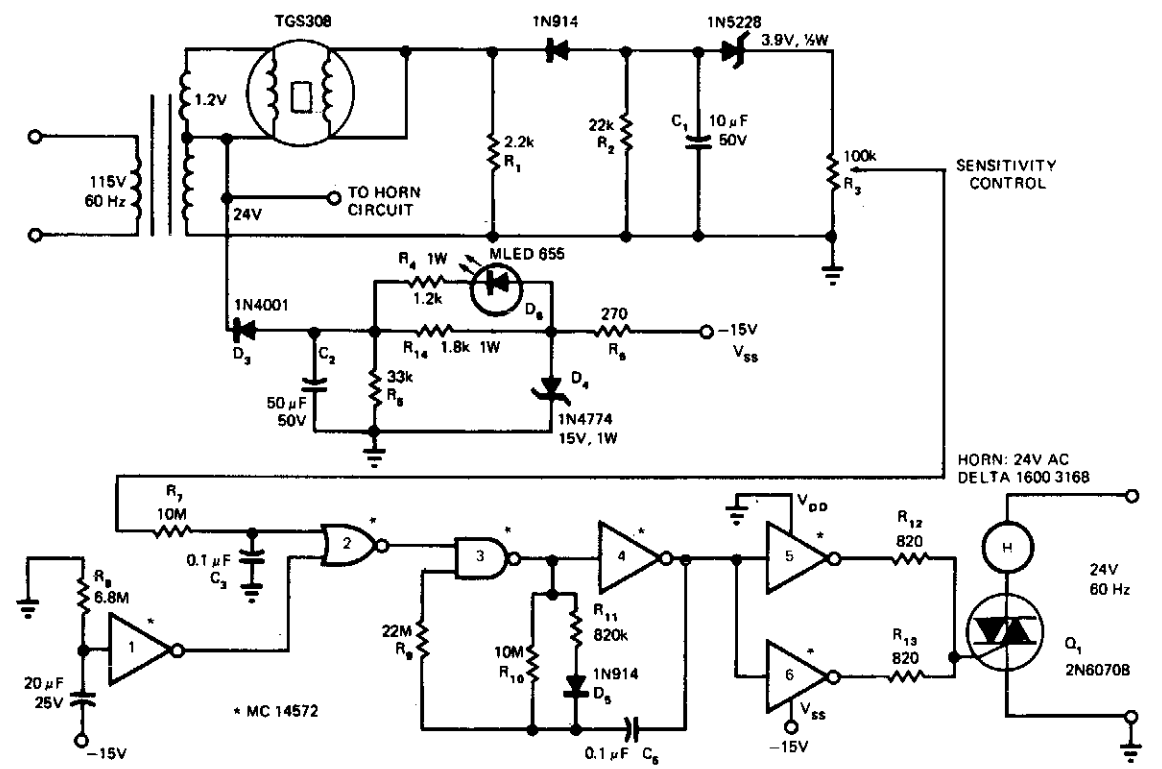
SCR SMOKE ALARM



POPULAR ELECTRONICS

Fig. 40-7

GAS/SMOKE DETECTOR



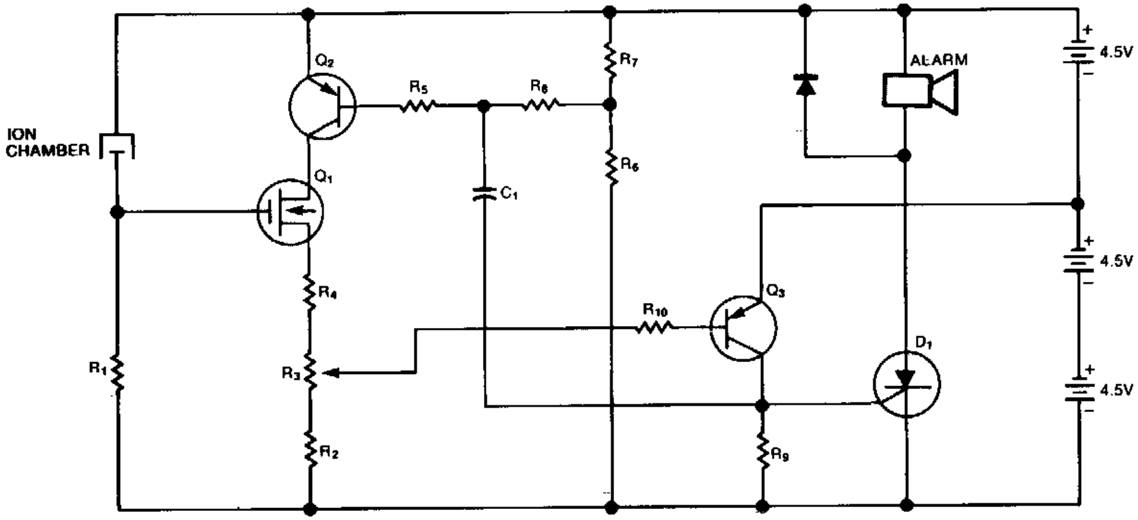
EDN

Fig. 40-8

In the presence of smoke or gas, the ac output voltage increases and becomes rectified, filtered and zener-diode coupled (D2 for thresholding) to sensitivity control R3. Under no gas condition, the output equals approximately 0 V (high). When gas is present, the output will be a negative value (low) sufficient to overcome the threshold of McMOS gate 2 and D2. The circuit shown uses a TGS 308 sensor, a general-purpose gas detector that is not sensitive to smoke or carbon monoxide. If smoke is the primary element to be detected, use the TGS 202 sensor. The two sensors are basically identical; the main differences lie in the heater voltage and the required warm up time delay. The TGS requires a 1.2 V heater and a 2 minute delay, whereas the TGS 202 requires 1.5 V and 5 minutes, respectively.

The system uses a McMOS gated oscillator directly interfacing with a triac-controlled ac horn. Using the MC14572 HEX functional gate, four inverters, one two-input NAND gate and one two-input NOR gate, the circuit provides the complete gas/smoke detector logic functions time delay, gated astable multi-vibrator control and buffers operation. The 24-Vac horn produces an 85/90-dB sound level output at a distance of 10 ft. Controlled by the astable multivibrator, the horn generates a pulsating alarm—a signal that may be advantageous over a continuous one in some noise environments.

SMOKE DETECTOR



EDN

Fig. 40-9

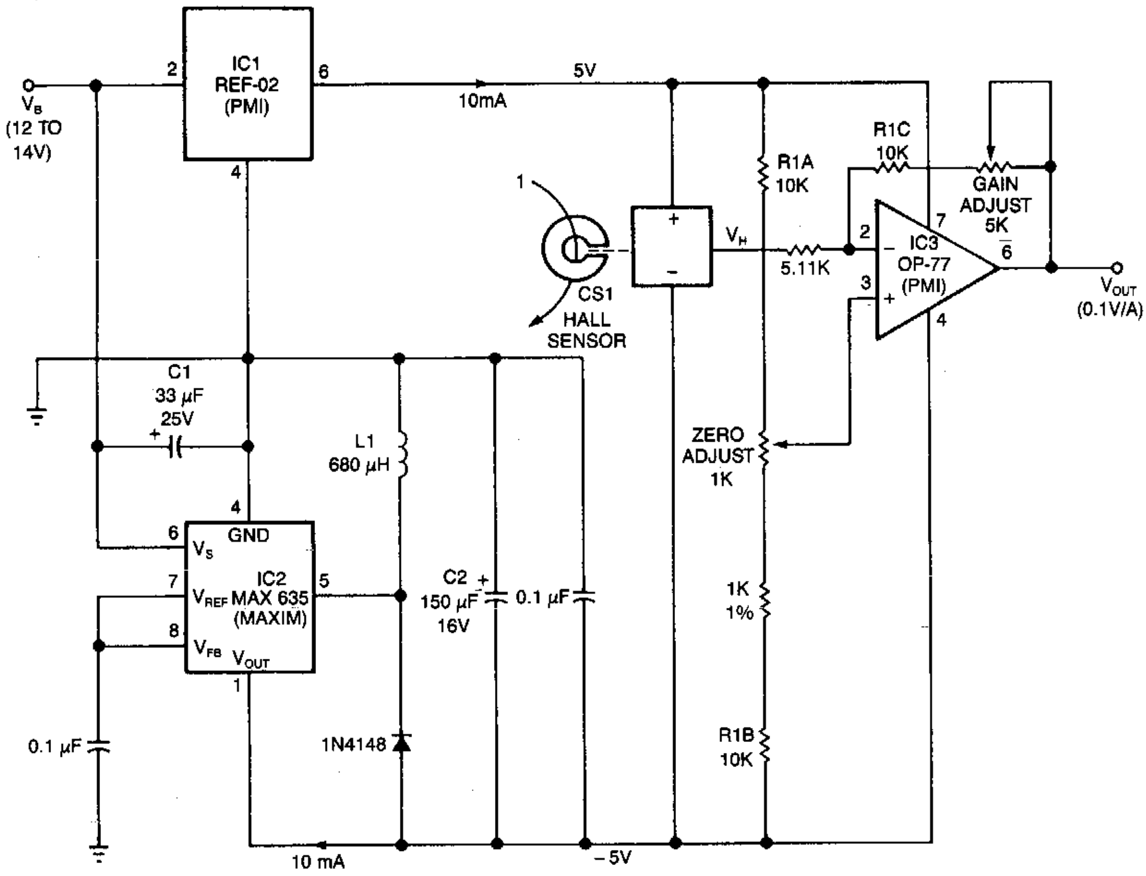
This circuit comes from U.S. Patent 3,778,800, granted to BRK Electronics in Aurora, IL. The circuit provides a smoke detector with an alarm for both smoke and low batteries. The R6/R7 voltage divider monitors the battery and will turn Q2 and Q1 off when the battery voltage falls too low. The smoke-detector chamber will also cut Q1 off when it senses smoke. Q1 via Q3, triggers SCR D1 and sounds the alarm. Capacitor C1 provides feedback that causes the alarm to sound intermittently. The smoke detector and low-battery circuits sound the alarm at two different rates.

41

Hall-Effect Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

CURRENT MONITOR



NOTES:
1. C1 AND C1 ARE 199D TANTALEX CAPACITORS FROM SPRAGUE
2. L1 IS A 6860-23 INDUCTOR FROM CADDELL-BURNS
3. R1A, R1B, AND R1C ARE PART OF A THIN-FILM RESISTOR NETWORK SUCH AS THE CADDOCK T914-10K.
4. CS1 IS A HALL-EFFECT CURRENT SENSOR (CSLA1CD) FROM MICROSWITCH.

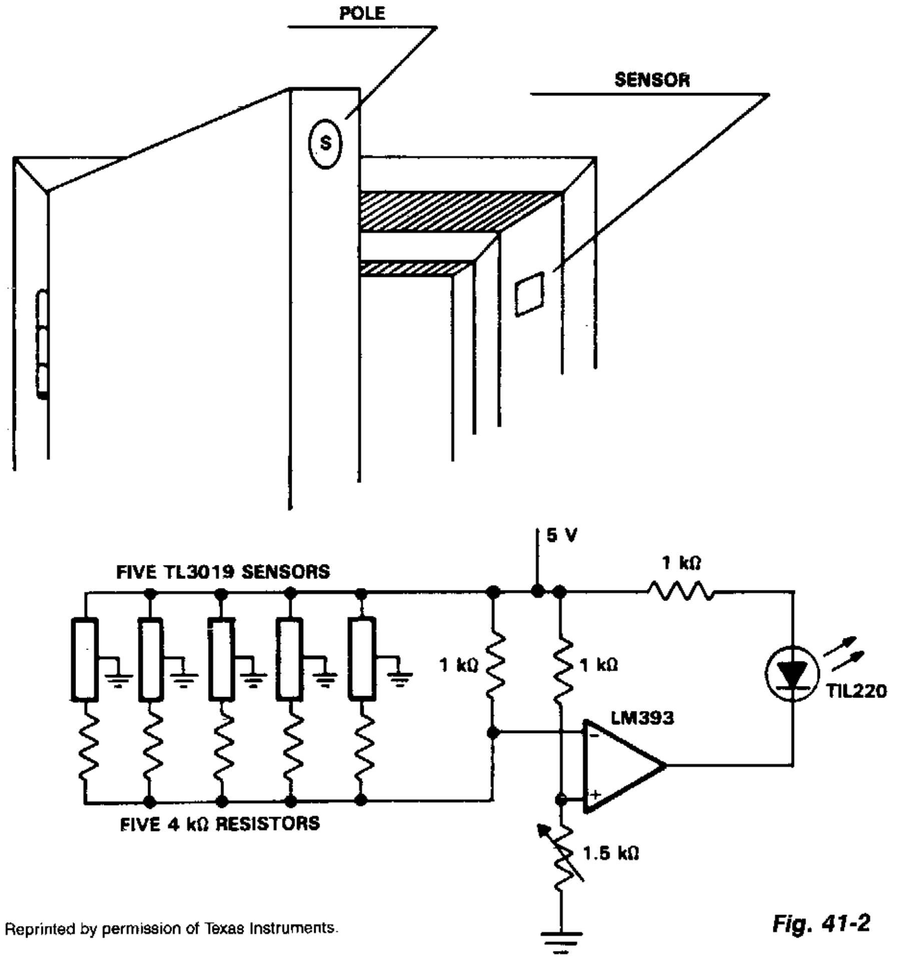
EDN

Fig. 41-1

This circuit uses a Hall-effect sensor, consisting of an IC that resides in a small gap in a flux-collector torrid, to measure dc current in the range of 0 to 40 A. You wrap the current-carrying wire through the toroid; the Hall voltage V_H is then linearly proportional to current I. The current drain from V_B is less than 30 mA.

To monitor an automobile alternator's output current, for example, connect the car battery between the circuit's V_B terminal and ground, and wrap one turn of wire through the toroid. Or, you could wrap 10 turns—if they fit—to measure 1 A full scale. When I = 0 V current sensor CS_I's V_H output equals one-half of its 10 V bias voltage. Because regulators IC1 and IC2 provide a bipolar bias voltage, V_H and V_{OUT} are zero when I is zero; you can then adjust the output gain and offset to scale V_{OUT} at 1 V per 10 A.

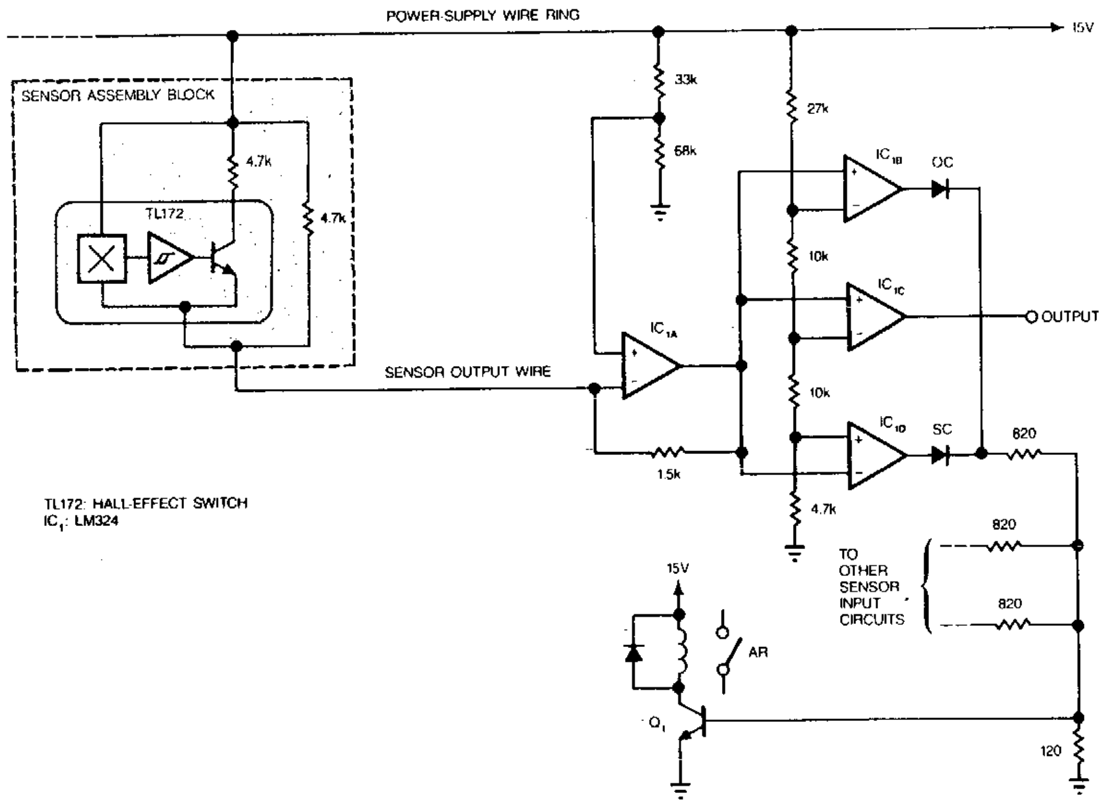
SECURITY DOOR AJAR ALARM



Reprinted by permission of Texas Instruments.

Fig. 41-2

In operation, the TL3019 device will activate, or become low, when a south pole of a magnet comes near the chip face of the device. The example shows five doors. Each door has a magnet embedded in its edge with the south pole facing the outer surface. At the point where the magnet is positioned with the door closed, a TL3019 sensor is placed in the door jamb. With the door closed, the Hall devices will be in a logic low state. This design has five doors and uses five TL3019 devices. Each TL3019 has a 4-KΩ resistor in series and all door sensor and resistor sets are in parallel and connected to the inverting input of an LM393 comparator. With all doors closed, the effective resistance will be about 800 Ω and produce 2.2 V at the inverting input. The noninverting input goes to a voltage divider network which sets the reference voltage. The 1.5-KΩ potentiometer is adjusted so the indicator goes out with all doors closed. This will cause 2.35 V to appear at the noninverting input of the comparator. When a door opens, the voltage at the inverting input will go to 2.5 V which is greater than V_{REF}, and the LED will light. A large number of doors and windows can be monitored with this type of circuit. Also, it could be expanded to add an audible alarm in addition to the visual LED.



EDN

Fig. 41-3

Hall-effect switches have several advantages over mechanical and optically coupled switches. They're insensitive to environmental light and dirt, they don't bind, and they don't sustain mechanical wear. Their major drawback is that they require three wires per device. The circuit shown, however, reduces this wire count to $N + 1$ wires for N devices.

Amplifier IC1A is configured as a current-to-voltage converter. It senses the sensor assembly's output current. When the Hall-effect switch is actuated, the sensor's output current increases to twice its quiescent value. Amplifier IC1B, configured as a comparator, detects this increase. The comparator's output decreases when the Hall-effect switch turns on.

The circuit also contains a fault-detection function. If any sensor output wire is open, its corresponding LED will turn on. If the power-supply line opens, several LEDs will turn on. A short circuit will also turn an LED on. Every time an LED turns on, Q1 turns on and the alarm relay is actuated.

HALL-EFFECT COMPASS

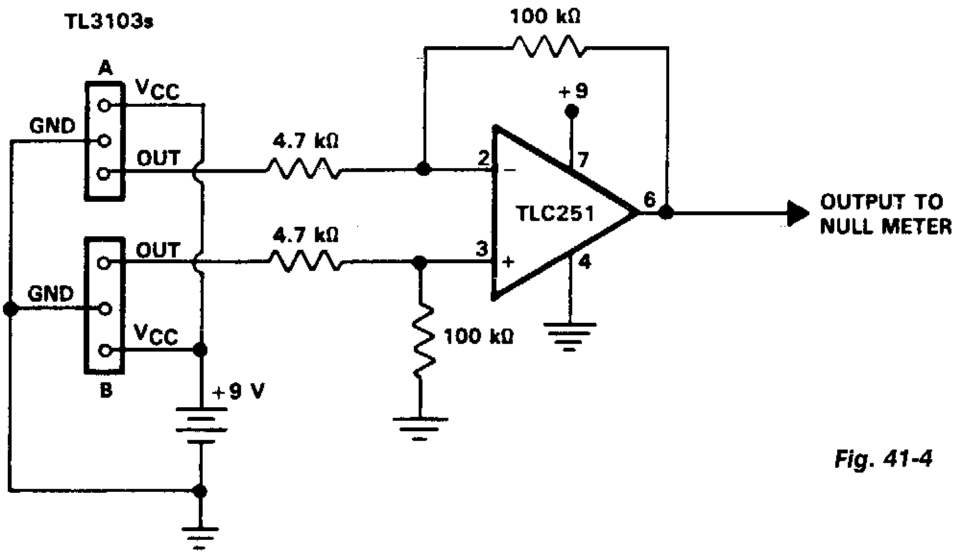


Fig. 41-4

Reprinted by permission of Texas Instruments.

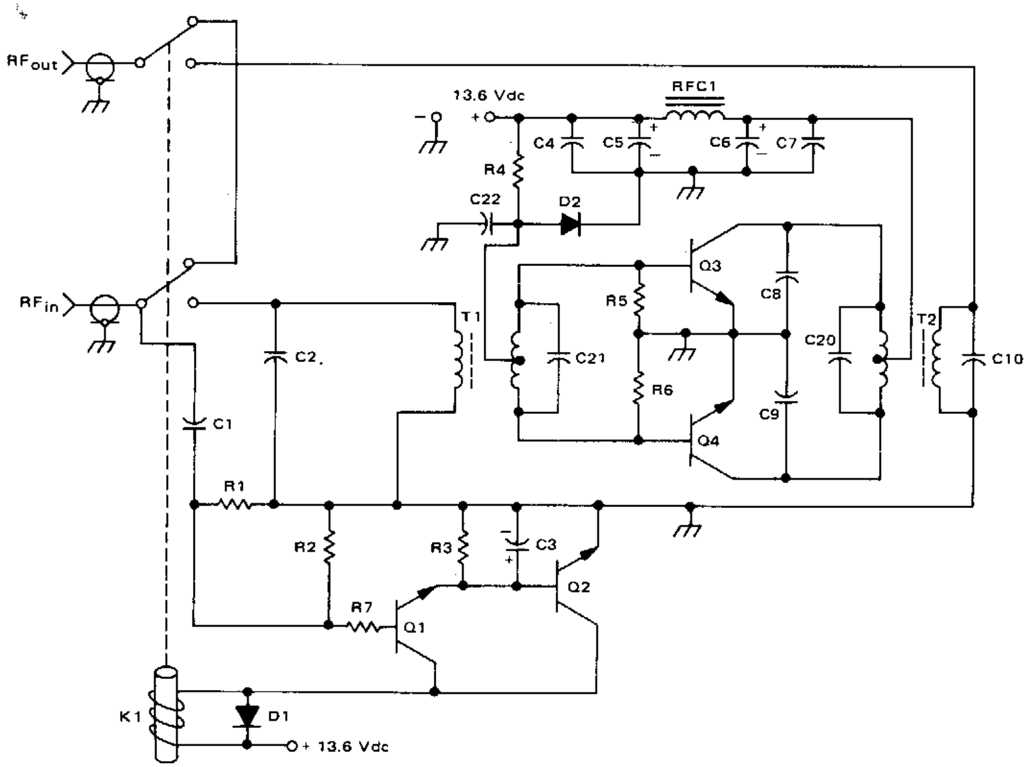
The TL3103 linear Hall-effect device can be used as a compass. By definition, the north pole of a magnet is the pole that is attracted by the magnetic north pole of the earth. The north pole of a magnet repels the north-seeking pole of a compass. By convention, lines of flux emanate from the north pole of a magnet and enter the south pole. The circuit of the compass is shown. By using two TL3103 devices instead of one, we achieve twice the sensitivity. With each device facing the opposite direction, device A would have a position output while the output of device B would be negative with respect to the zero magnetic field level. This gives a differential signal to apply to the TLC251 op amp. The op amp is connected as a difference amplifier with a gain of 20. Its output is applied to a null meter or a bridge balance indicator circuit.

42

High-Frequency Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

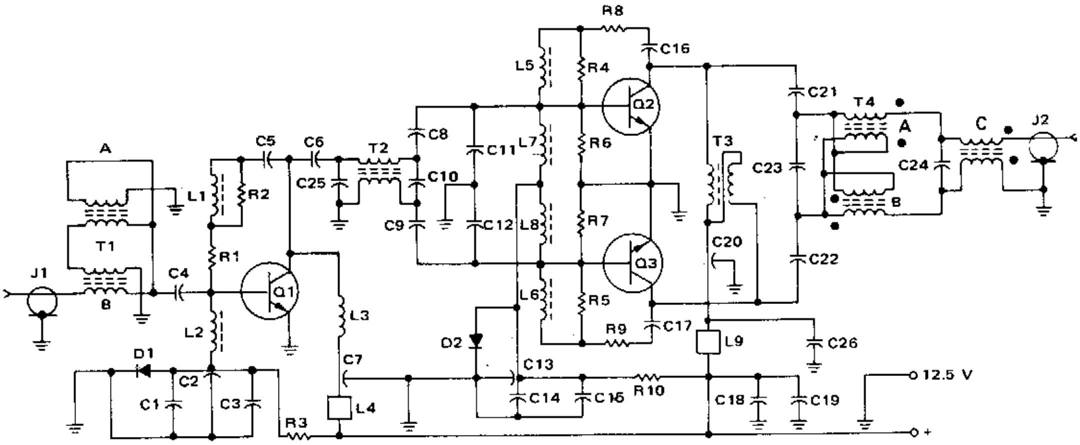
2 – 30 MHz 140-W (PEP) AMATEUR RADIO LINEAR AMPLIFIER



C1	= 33 pF Dipped Mica	R7	= 100 Ω 1/4 W Resistor
C2	= 18 pF Dipped Mica	RFC1	= 9 Ferroxcube Beads on #18 AWG Wire
C3	= 10 μF 35 Vdc for AM operation, 100 μF 35 Vdc for SSB operation.	D1	= 1N4001
		D2	= 1N4997
C4	= .1 μF Erie	Q1, Q2	= 2N4401
C5	= 10 μF 35 Vdc Electrolytic	Q3, 4	= MRF454
C6	= 1 μF Tantalum	T1, T2	= 16:1 Transformers
C7	= .001 μF Erie Disc	C20	= 910 pF Dipped Mica
C8, 9	= 330 pF Dipped Mica	C21	= 1100 pF Dipped Mica
R1	= 100 kΩ 1/4 W Resistor	C10	= 24 pF Dipped Mica
R2, 3	= 10 kΩ 1/4 W Resistor	C22	= 500 μF 3 Vdc Electrolytic
R4	= 33 Ω 5 W Wire Wound Resistor	K1	= Potter & Brumfield KT11A 12 Vdc Relay or Equivalent
R5, 6	= 10 Ω 1/2 W Resistor		

Fig. 42-1

The amplifier operates across the 2 – 30 MHz band with relatively flat gain response and reaches gain saturation at approximately 210 W of output power. Both input and output transformers are 4:1 turns ratio (16:1 impedance ratio) to achieve low input SWR across the specified band and a high saturation capability. When using this design, it is important to interconnect the ground plane on the bottom of the board to the top, especially at the emitters of the MRF454s.



C1, C14, C18 – 0.1 μF ceramic.
C2, C7, C13, C20 – 0.001 μF feed through.
C3 – 100 μF/3V.
C4, C6 – 0.033 μF mylar
C5 – 0.0047 μF mylar.
C8, C9 – 0.015 and 0.033 μF mylars in parallel.
C10 – 470 pF mica.
C11, C12 – 560 pF mica.
C15 – 1000 μF/3 V
C16, C17 – 0.015 μF mylar
C19 – 10 pF 15 V
C21, C22 – two 0.068 μF mylars in parallel.
C23 – 330 pF mica
C24 – 39 pF mica
C25 – 680 pF mica
C26 – .01 μF ceramic

R1, R6, R7 – 10 Ω, 1/2 W carbon.
R2 – 51 Ω, 1/2 W carbon
R3 – 240 Ω, 1 wire W
R4, R5 – 18 Ω, 1 W carbon
R8, R9 – 27 Ω, 2 W carbon
R10 – 33 Ω, 6 W wire W

L1 – 0.22 μh molded choke
L2, L7, L8 – 10 μh molded choke
L5, L6 – 0.15 μh
L3 – 25 t, #26 wire, wound on a 100 Ω, 2 W resistor. (1.0 μh)
L4, L9 – 3 ferrite beads each.

T1 – 2 twisted pairs of #26 wire, 8 twists per inch. A = 4 turns, B = 8 turns. Core- -Stackpole 57-9322-11, Indiana General F627-8Q1 or equivalent
T2 – 2 twisted pairs of #24 wire, 8 twists per inch, 6 turns. (Core as above.)
T3 – 2 twisted pairs of #20 wire, 6 twists per inch, 4 turns. (Core as above.)
T4 – A and B = 2 twisted pairs of #24 wire, 8 twists per inch, 5 turns each. C = 1 twisted pair of #24 wire, 8 turns. Core - -Stackpole 57-9074-11, Indiana General F624-19Q1 or equivalent.

Q1 – 2N6367
Q2, Q3 – 2N6368

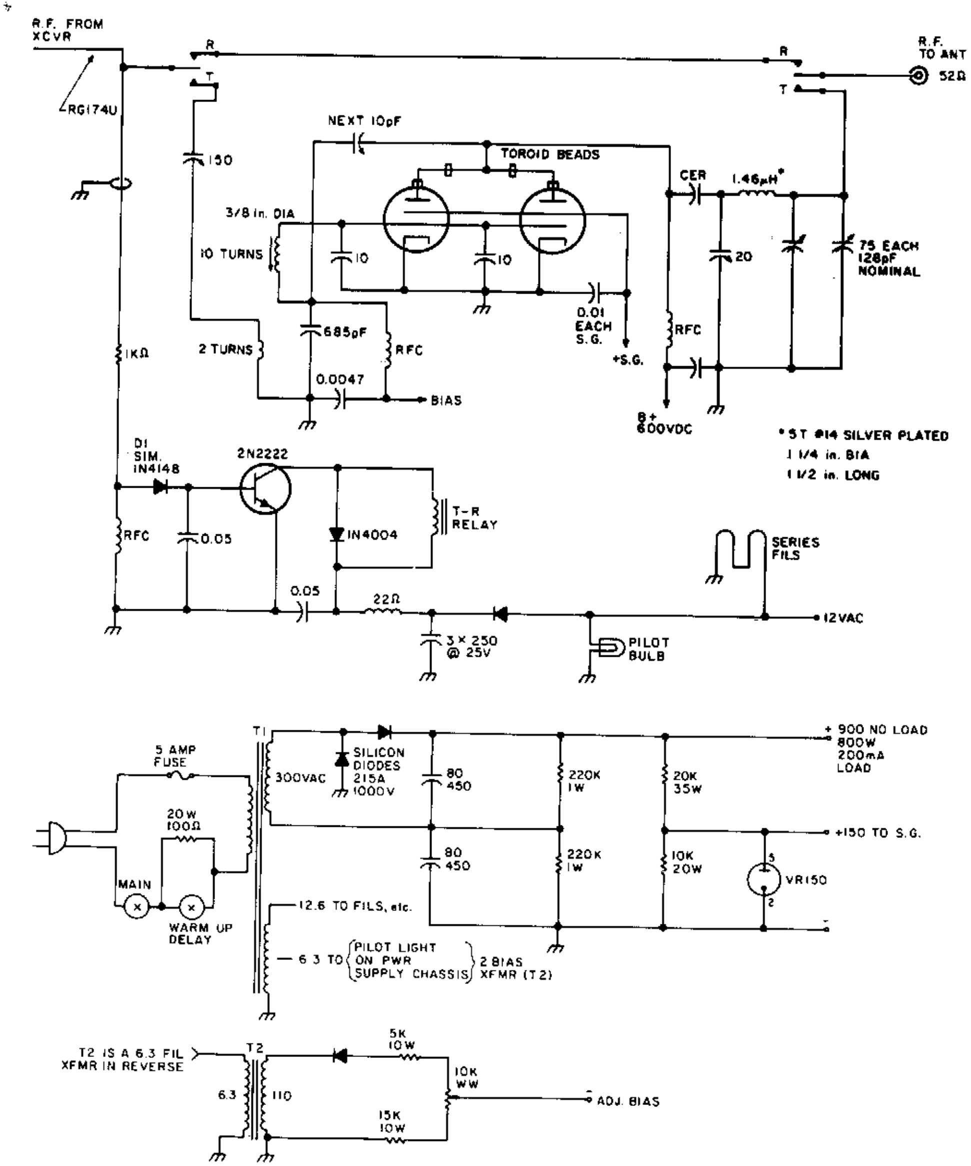
D1 – 1N4001
D2 – 1N4997

J1, J2 – BNC connectors

Fig. 42-2

This amplifier utilizes a 2N6367 and a pair of 2N6368 transistors. The 2N6367 transistor is employed as a driver and is specified for up to 9 W (PEP) output. In the amplifier design the driver must supply on 5 W (PEP) at 30 MHz with a resulting IMD performance of about –37 to –38 dB. At lower operating frequencies, drive requirements drop to the 2–3 W (PEP) range and IMD performance improves to better than 40 dB. Two 2N6368 transistors are employed in the final stage of the transmitter design in a push-pull configuration. These devices are rated at 40 W (PEP) and –30 dB maximum IMD, although –35 dB performance is more typical for narrowband operation. Without frequency compensation, the completed amplifier can deliver 90 W (PEP) in the 25–30 MHz band with IMD performance down –30 dB. If only the power amplifier stage is frequency compensated, 95 W (PEP) can be obtained at 6–10 MHz.

29-MHz AMPLIFIER

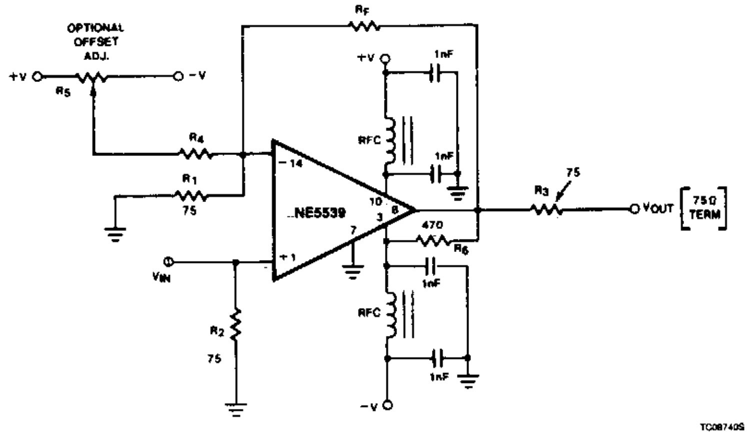


73 AMATEUR RADIO

Fig. 42-3

The only adjustments that require close attention are input, output, and neutralization. The 150-pF capacitor in the input line compensates for impedance mismatch. You tune for maximum signal transfer from exiter to final with an in-line meter or external field strength meter. The final is a conventional pi network. When neutralized, the plate current dip should be at about the same setting of the 20-pF plate capacitor as maximum output. Adjust bias to let tubes idle at about 30 mA.

28-dB NONINVERTING AMPLIFIER



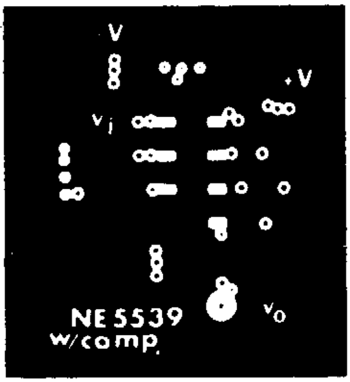
NOTES:

R_1 = 75Ω 5% CARBON
R_2 = 75Ω 5% CARBON
R_3 = 75Ω 5% CARBON
R_4 = 36k 5% CARBON

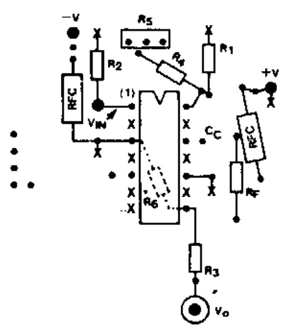
R_5 = 20k TRIMPOT (CERMET)
R_F = 1.5k (28dB GAIN)
R_6 = 470Ω 5% CARBON

RFC 3T # 26 BUSS WIRE ON
FERROXCUBE VK 200 09/3B CORE
BYPASS CAPACITORS
1nF CERAMIC
(MEPCO OR EQUIV.)

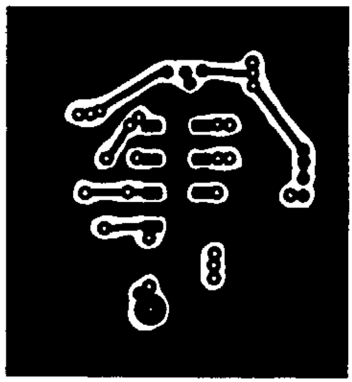
Top Plane Copper[1] (Component Side)



Component Side (Component Layout)



Bottom Plane Copper[1]



NOTES:
(X) indicates ground connection to top plane.
*R_6 is on bottom side.

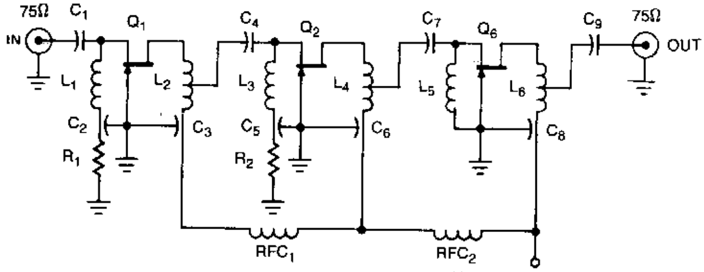
NOTE:
1. Bond edges of top and bottom ground plane copper.

SIGNETICS

Fig. 42-4

The physical circuit layout is extremely critical. Breadboarding is not recommended. A double-sided copper-clad printed circuit board will result in more favorable system operation.

WIDEBAND UHF AMPLIFIER WITH HIGH-PERFORMANCE FETS



C_1, C_4, C_7, C_9 = 68 pF
C_2, C_5 = 500 pF
C_3, C_6, C_8 = 1000 pF
Q_1, Q_2, Q_3 = Siliconix U310
L_1, L_3, L_5 = 120 nHy
L_2, L_4, L_6 = 222 nHy
RFC_1, RFC_2 = 2.2 nHy
R_1, R_2 = 51Ω
V_D = +20V

SILICONIX

Fig. 42-5

The amplifier circuit is designed for a 225 MHz center frequency, 1 dB bandwidth of 50 MHz, low-input VSWR in a 75-Ω system, and 24 dB gain. Three stages of U310 FETs are used, in a straight-forward design.

BROADCAST BAND RF AMPLIFIER

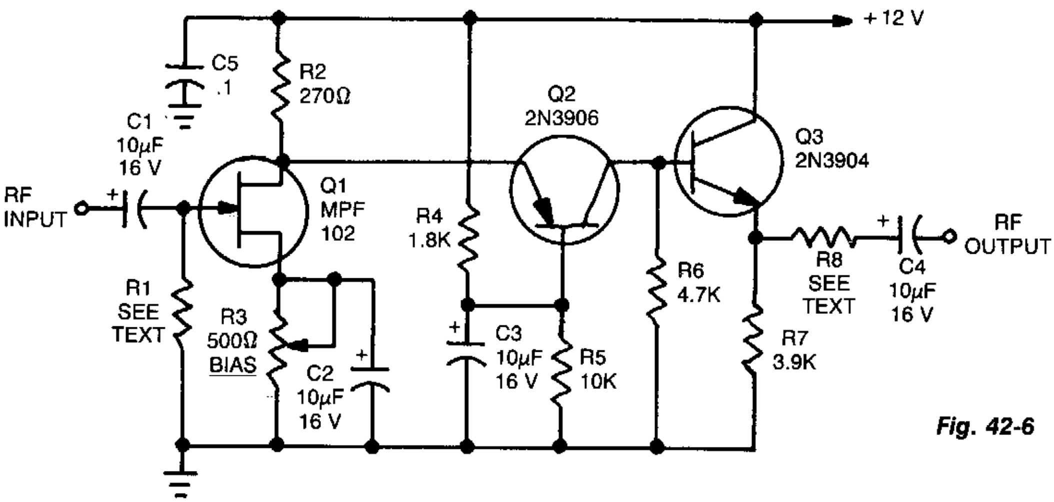


Fig. 42-6

The circuit has a frequency response ranging from 100 Hz to 3 MHz; gain is about 30 dB. Field-effect transistor Q1 is configured in the common-source self-biased mode. Optional resistor R1 allows you to set the input impedance to any desired value; commonly, it will be 50 Ω.

The signal is then direct coupled to Q2, a common-base circuit that isolates the input and output stages and provides the amplifier's exceptional stability. Last, Q3 functions as an emitter follower, to provide low output impedance at about 50 Ω. If you need higher output impedance, include resistor R8. It will affect impedance according to this formula: $R8 \approx R_{OUT} - 50$. Otherwise, connect output capacitor C4 directly to the emitter of Q3.

MINIATURE WIDEBAND AMPLIFIER

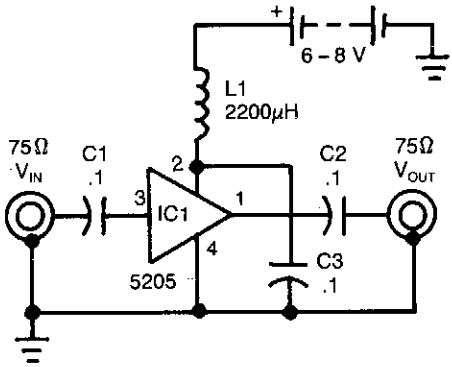
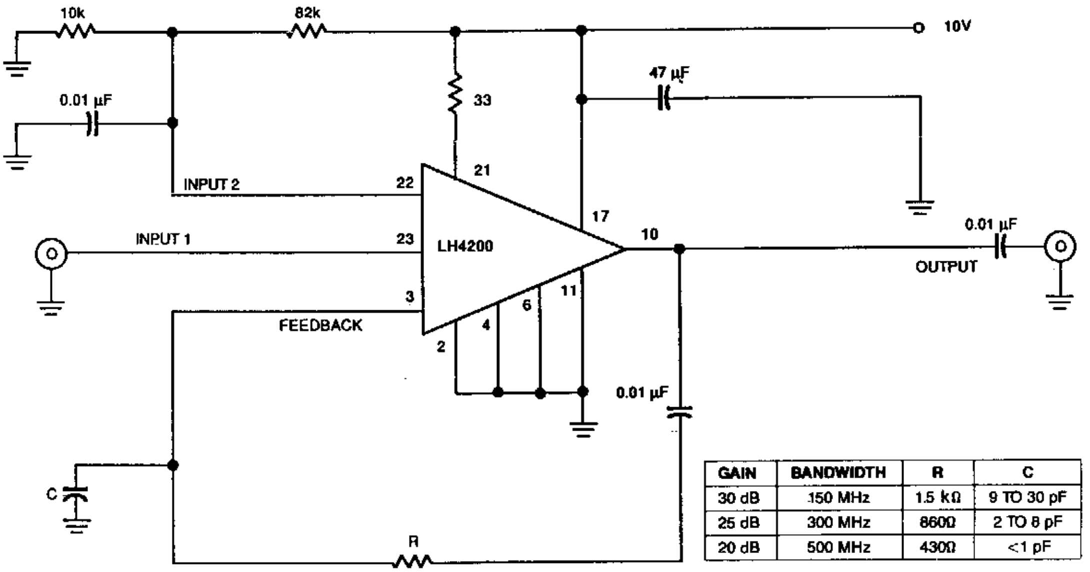


Fig. 42-7

This wideband amplifier uses only five components. External signals enter pin 3 of IC1 via ac coupling capacitor C1. Following amplification, the boosted signals from IC1 pin 1 are coupled to the output by capacitor C2. Capacitor C3 decouples the dc power supply, while rf current is isolated from the power supply by rf choke L1.

The NE5205's low current consumption of 25 mA at 6 Vdc makes battery-powered operation a reality. Although the device is rated for a 6 to 8 V power supply, 6 V is recommended for normal operation. From 6 V an internal bias of 3.3 V results, which permits a 1.4 V pk-pk output swing for video applications.

WIDEBAND 500 kHz – 1 GHz HYBRID AMPLIFIER



GAIN	BANDWIDTH	R	C
30 dB	150 MHz	1.5 kΩ	9 TO 30 pF
25 dB	300 MHz	860Ω	2 TO 8 pF
20 dB	500 MHz	430Ω	<1 pF

EDN

Fig. 42-8

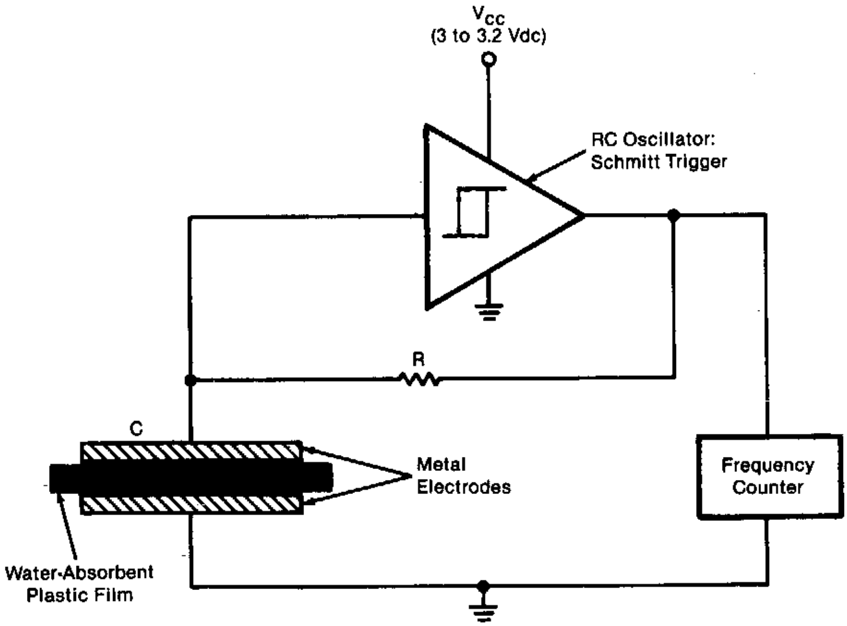
The amplifier's input stage is a dual-gate GaAs FET, which provides low input capacitance and high transconductance. The dual-gate structure accepts the signal on input 1. Input 2 controls the gain of the amplifier. The amplifier has a third input for use in series feedback. The output feeds back to pin 3 via a single resistor, which controls the overall power gain of the amplifier. At 10 MHz, the output is capable of delivering 12 dBm into a 50-Ω load with 1 dB of signal compression. The ac-coupled amplifier has a gain of 37 dB at 100 MHz and 3 dB at 1 GHz.

43

Humidity Sensor

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

LOW-COST HUMIDITY SENSOR



NASA TECH BRIEFS

Fig. 43-1

The sensor is an RC oscillator in which a water-absorbent plastic film is the insulator in the capacitive element. The capacitance of the film increases with the amount of water it absorbs from the air, and thus reduces the oscillation frequency. A frequency counter produces a digital output that represents the change in frequency and hence the change in relative humidity. The sensor can be used to measure humidity in the atmosphere, in the soil, and in industrial gases, for example. A Schmitt-trigger-type IC is connected to the capacitor, which consists of a film of a commercially produced sulfonated fluorocarbon polymer, 2 in. (5.08 cm) square, sandwiched between perforated metal plates. The oscillation frequency decreases almost linearly from about 100 to 16 kHz as the relative humidity increases from about 20 to 76%.

44

Indicators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

STEREO INDICATOR

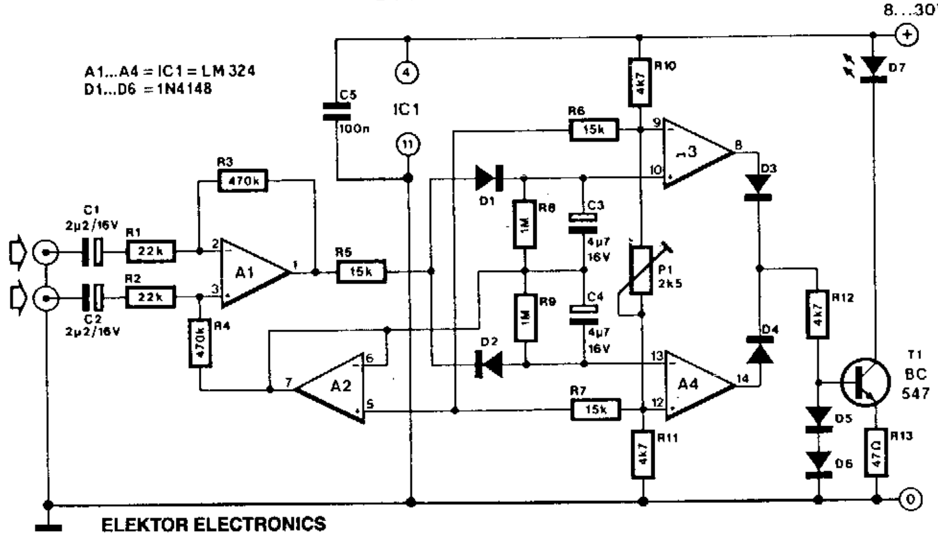


Fig. 44-1

On most FM tuners, the stereo indicator lights upon detection of the 19-kHz pilot tone. However, this doesn't mean that the program is actually stereophonic, since the pilot tone is often transmitted with mono programs also. A similar situation exists on stereo amplifiers, where the stereo LED is simply controlled from the mono/stereo switch.

The LED-based stereo indicator described here lights only when a true stereo signal is fed to the inputs. Differential amplifier A1 raises the difference between the L and R input signals. When these are equal, the output of A1 remains at the same potential as the output of A2, which forms a virtual ground rail at half the supply voltage. When A1 detects a difference between the L and R input signals, it supplies a positive or negative voltage with respect to the virtual ground rail, and so causes C3 to be charged via D1 or C4 via D2. Comparator A3/A4 switches on the LED driver via OR circuit D3/D4. The input signal level should not be less than 100 mV to compensate for the drop across D1 or D2. The sensitivity of the stereo indicator is adjustable with P1.

VISUAL LEVEL INDICATOR

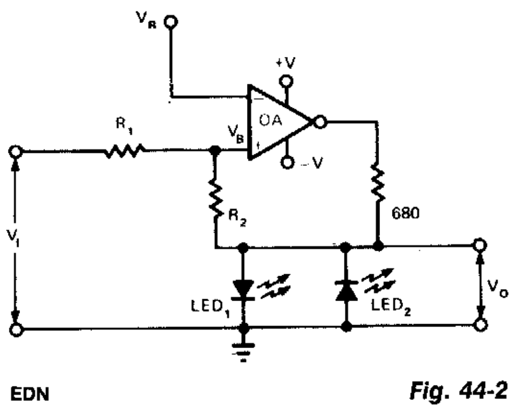
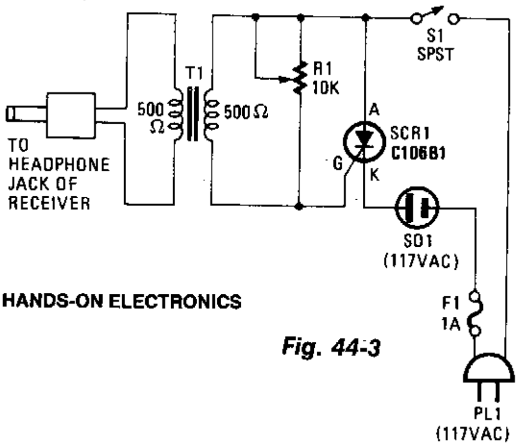


Fig. 44-2

This indicator is basically a switch with hysteresis characteristics. If the input voltage momentarily (or permanently) exceeds the most positive reference level, LED1 is switched on. If, on the other hand, the voltage falls below the negative, or least positive, reference level, LED1 will be switched off and LED2 switched on. The output voltage, V_O is clamped either to the diode voltage V_{D1}, or V_{D2} depending on which LED is conducting. For V_O to be positive, V_B has to be positive with respect to the reference voltage V_R; for V_O to be negative, V_B has to be negative with respect to V_R.

RECEIVER SIGNAL ALARM



HANDS-ON ELECTRONICS

Fig. 44-3

RF-ACTUATED RELAY

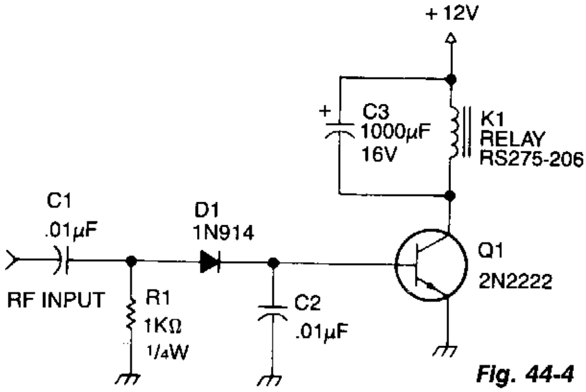
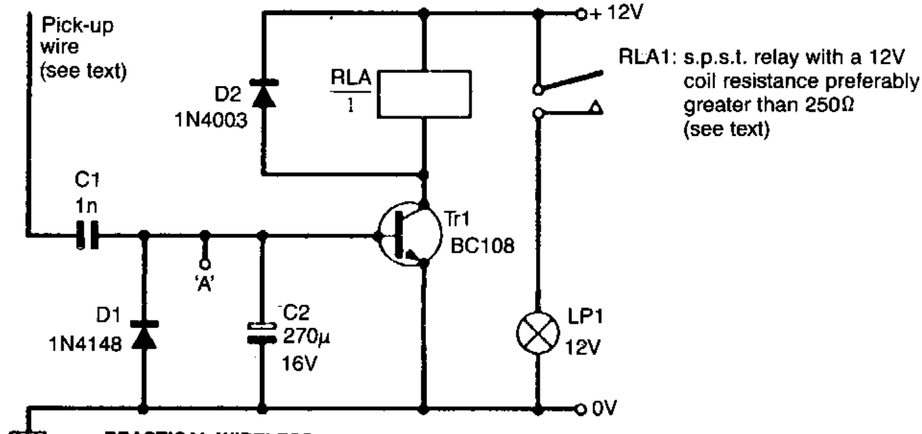


Fig. 44-4

73 MAGAZINE

Automatic antenna switching or rf power indication can be achieved with this circuit. Relay will key with less than 150 mW drive on 2 m.

ON-THE-AIR INDICATOR



PRACTICAL WIRELESS

Fig. 44-5

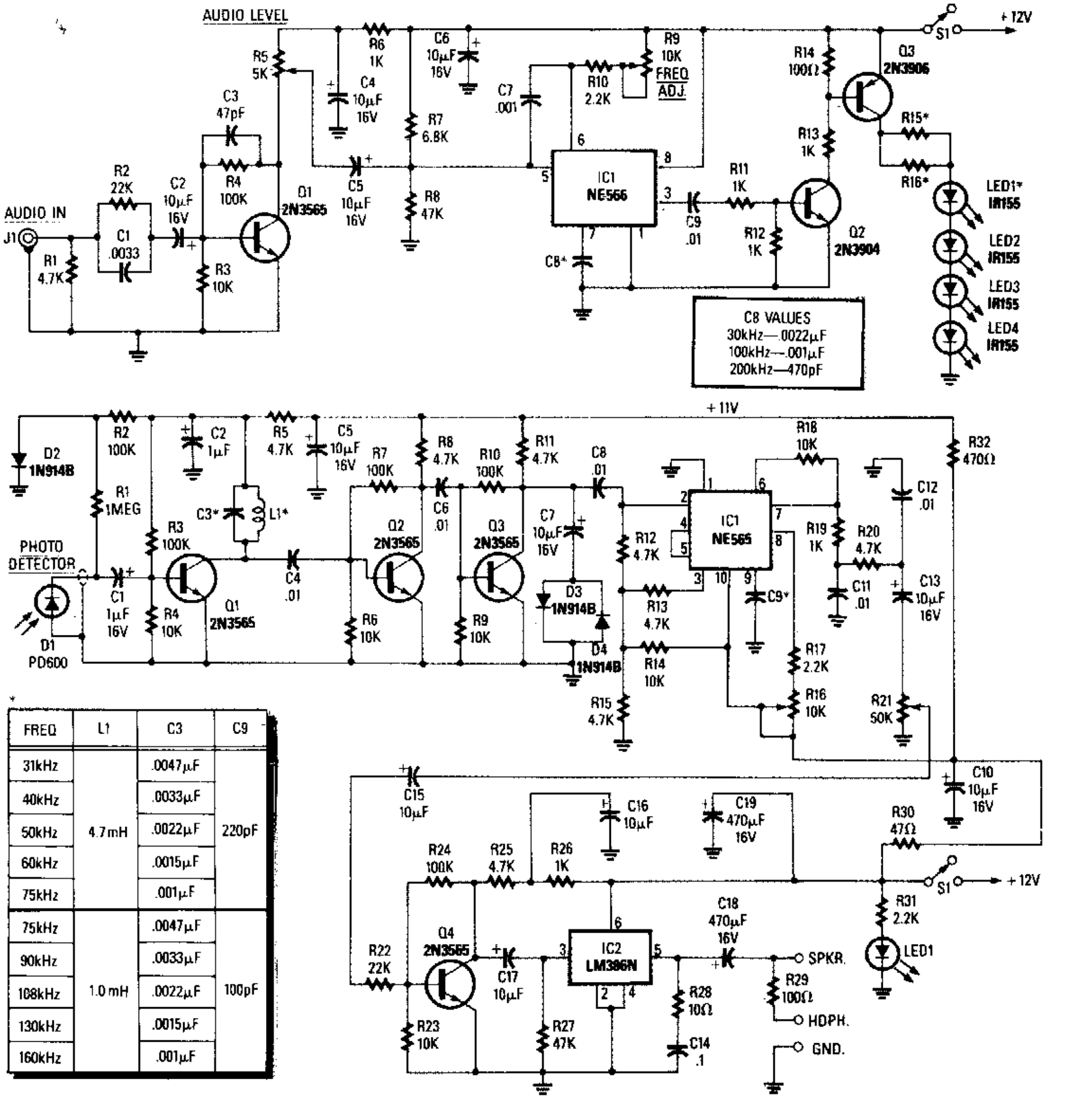
The circuit is a simple rf-actuated switch which will respond to any strong field in the region of the pickup wire. The length of the wire will depend on how much coupling is needed, but a 250-mm length wrapped around the outside of the coaxial cable feeding the antenna should suffice for most power levels. If only one band is used, the wire can be made a resonant length—495 mm for 144 MHz band operation for example. When rf energy is picked up by the device, diode D1 will conduct on the negative half-cycles, but will be cut off on the positive half-cycles. The result will be a net positive voltage at the base of transistor Tr1, forward biasing it into conduction. On ssb and cw transmissions, where the transmission is not continuous, that bias would be constantly varying and the relay RLA would chatter. However, capacitor C2 holds the bias voltage steady until a long gap in transmissions occurs.

45

Infrared Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

INFRARED WIRELESS SPEAKER SYSTEM



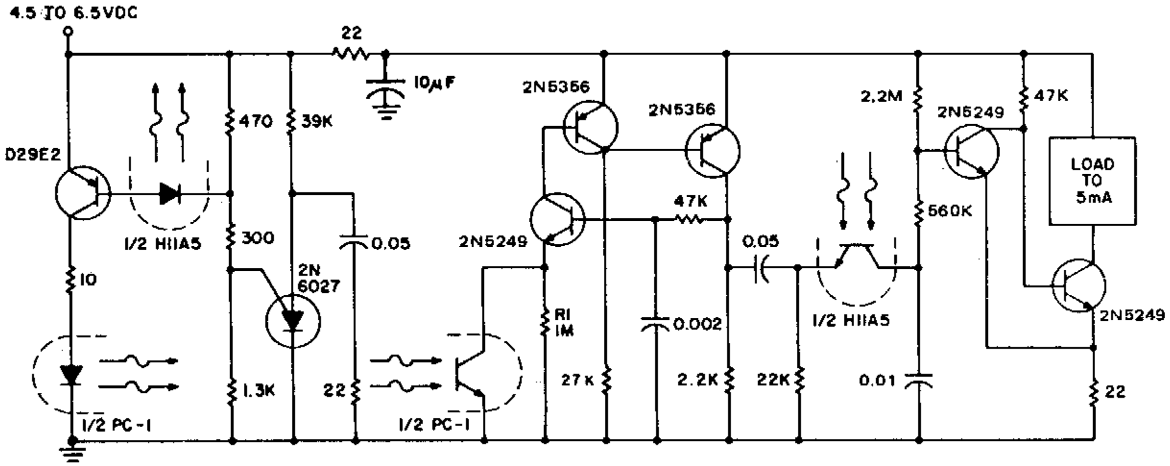
FREQ	L1	C3	C9
31kHz	4.7mH	.0047μF	220pF
40kHz		.0033μF	
50kHz		.0022μF	
60kHz		.0015μF	
75kHz		.001μF	
75kHz	1.0mH	.0047μF	100pF
90kHz		.0033μF	
108kHz		.0022μF	
130kHz		.0015μF	
160kHz		.001μF	

Fig. 45-1

Although the IR region is free from radio interference, it is subject to interference from incandescent lamps, fluorescent lamps, stray reflections, and other sources.

A simple way to overcome that problem is to create a *carrier* by chopping the IR radiation at a rate of 100 kHz. The audio then modulates the carrier by modulating the chopping rate. A receiver then detects the IR beam as a 100-kHz FM signal. The only disadvantage is that instead of a simple audio amplifier, a high-gain FM receiver is necessary. However, with the ICs that are now available, an FM receiver is easy to build, and contains little more circuitry than a high-gain audio amplifier. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

LONG-RANGE OBJECT DETECTOR



PC-1 SELECTION	TRANSMISSION RANGE	REFLECTIVE RANGE
H23A1	5″	1″
LED56 and L14Q1	12″	3″
LED56 and L14G1	18″	4½″
LED55C and L14G1	32″	8″
1N6266 and L14G3	48″	12″
F5D1 and L14G3	80″	20″
F5D1 and L14P2	200″	50″

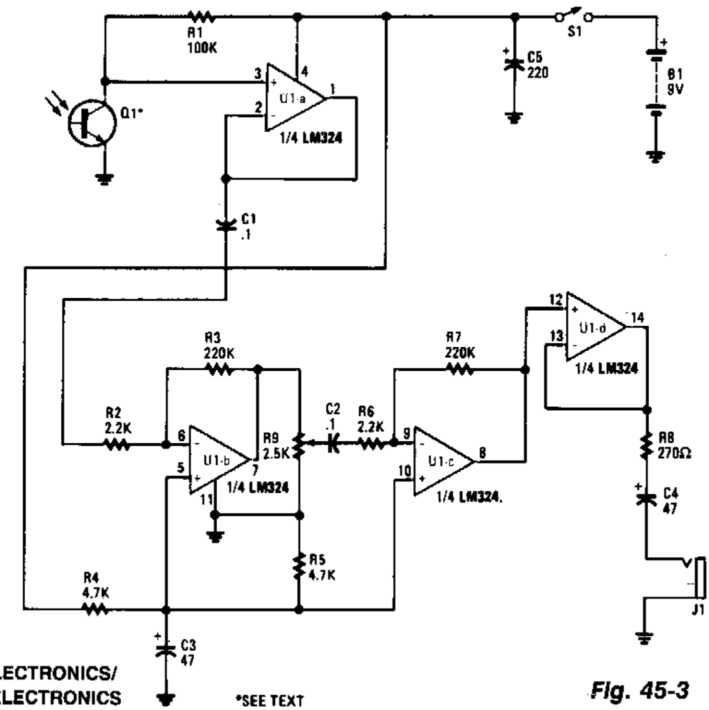
GE

When long ranges must be worked with IR light sources, and when high system reliability is required, pulsed-mode operation of the IR is required. Additional reliability of operation is attained by synchronously detecting the photodetector current, as this circuit does. PC-1 is an IR and phototransistor pair which detect the presence of an object blocking the transmission of light from the IR to the phototransistor. Relatively long-distance transmission is obtained by pulsing the IR, with about 10-μs pulses, at a 2-ms period, to 350 mA via the 2N6027 oscillator. The phototransistor current is amplified by the 2N5249 and 2N5356 amplifier to further increase distance and allows use of the H11A5, also pulsed by the 2N6027, as a synchronous detector, providing a fail-safe, noise immune signal to the 2N5249 pair forming a Schmitt-trigger output.

This design was built for battery operation, with long battery life a primary consideration. Note that another stage of amplification driving the IR can boost the range limited by the IR V_F, by 5 to 10 times. A higher supply voltage for the IR can double this range.

Today, optoelectronics are mostly used to transmit electronic information over light beams. These applications range from the use of optocouplers transmitting information between IC logic circuits and power circuits, between power lines and signal circuits, between telephone lines and control circuitry, to the pulse-modulated systems which transmit information through air or fiber optics over relatively great distances.

IR RECEIVER

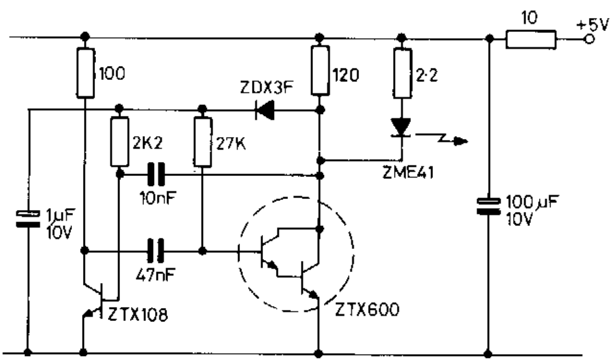


POPULAR ELECTRONICS/
HANDS-ON ELECTRONICS

Fig. 45-3

Infrared emissions detected by Q1 are fed through U1a to U1b, which amplifies the signal by a factor of 100. The amplified output of U1b is fed to U1c through R9, C2, and R6. Potentiometer R9 serves as a volume control. With R9 set to pass the maximum signal, U1c provides a gain of 100, for a total system gain of 10,000 dB. The output of U1c is connected to voltage follower circuit U1d to better match and drive headphones that can be plugged into J1.

IR TRANSMITTER

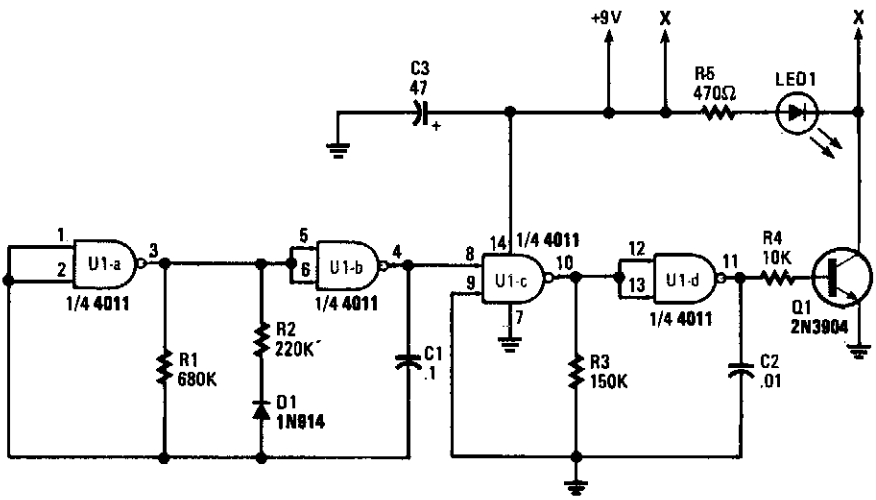


ZETEX, formerly FERRANTI

Fig. 45-4

The transmitter consists of an oscillator which drives a high output IR emitting diode. The oscillator is a sure start multivibrator circuit that provides an output of 15 to 1000 mark to space ratio at a frequency of 1 kHz. This large mark to space ratio allows the IR diode to be operated at a high peak current, provided by the ZTX600 Darlington transistor, to maximize the transmitter range. A decoupling network is included in the power supply lead to isolate it from any logic circuitry using the same 5-V power supply source. The transmitter supply current is approximately 65 mA.

DIGITAL IR TRANSMITTER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 45-5*

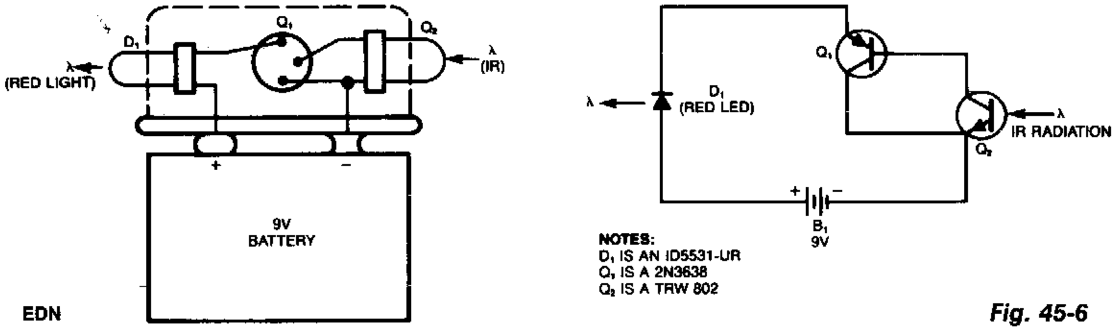
Gates U1a and U1b are configured as a low-frequency oscillator. The output waveform at pin 11 is nonsymmetrical with the positive portion of the signal, making up only 20% of the time period.

Diode D1, a 1N914 general-purpose unit, together with C1, R1, and R2, determine the on time for the positive portion of the output waveform. The off, or negative portion of the output waveform, depends mainly on the values of R1 and C1. The operating frequency of that oscillator is about 11 Hz. The second oscillator consists of U1c and U1d, which outputs on almost symmetrical waveform at a frequency of about 400 Hz. The output of first oscillator U1a/U1b is fed to pin 8 of U1c to key second oscillator U1c/U1d on and off at about 11 Hz, with the on time limited to about 20% of the time period (about 15 ms).

The output waveform of the second oscillator is fed to the base of Q1, which is used to drive IR diode LED1 in short bursts. Pulsing LED1 helps to save battery power, and also allows each circuit to be given its own special sound footprint.

By changing any of the values of R1, R2, R3, C1, or C2, the sound footprint can be varied. As the component values are increased, the oscillator's frequency goes down, and as the values are decreased, the frequency goes up.

SIMPLE IR DETECTOR

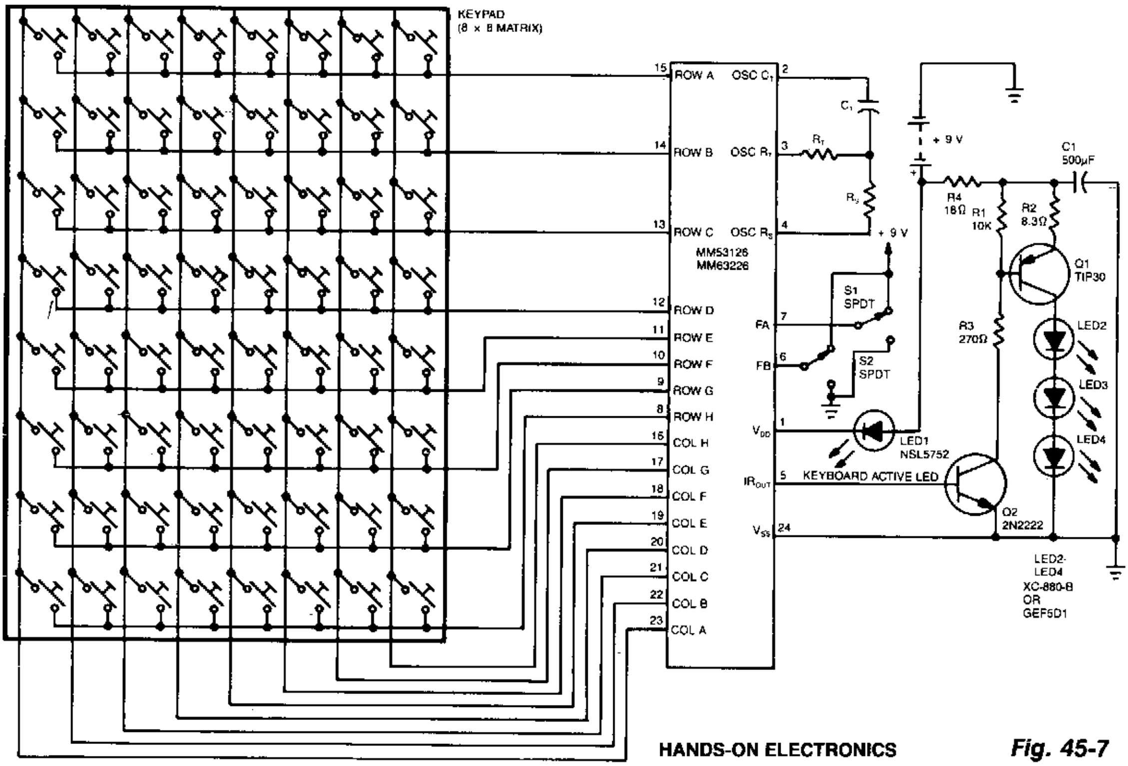


EDN

Fig. 45-6

This simple IR detector turns on a real LED when Q2 is exposed to invisible IR radiation found in fiber-optics systems, position sensors, and TV remote-control units. The device can be built on top of a 9-V battery and held in place with RTV. Its power dissipation is virtually zero, unless IR radiation or high ambient light is present. Normal fluorescent lighting is not a problem, but if necessary add an IR filter to the Q2 detector to exclude ambient light. Exposing the detector to a strong light or IR source gives a quick check of the battery and the red LED.

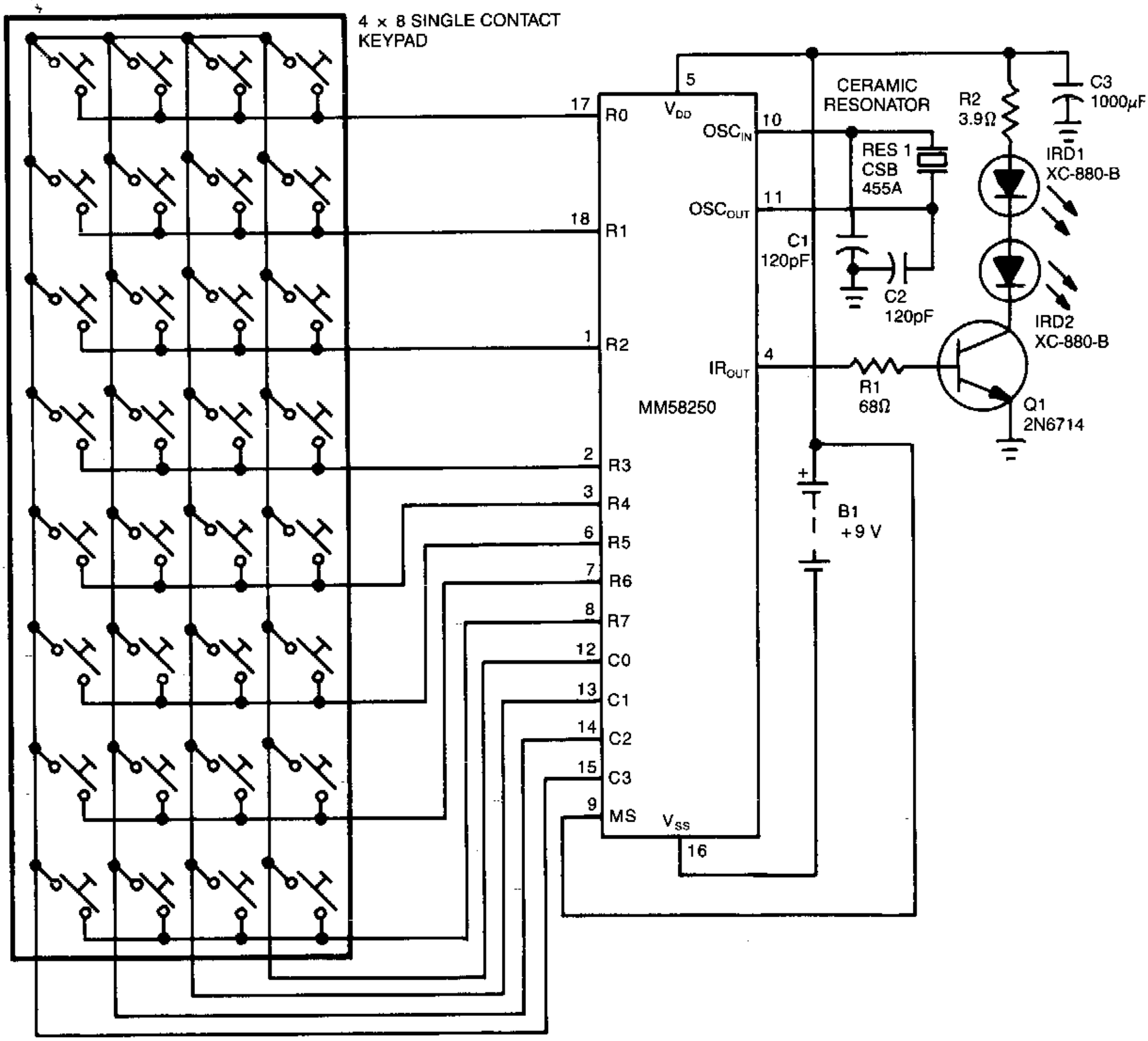
INFRARED TRANSMITTER



HANDS-ON ELECTRONICS

Fig. 45-7

INFRARED TRANSMITTER



POPULAR ELECTRONICS

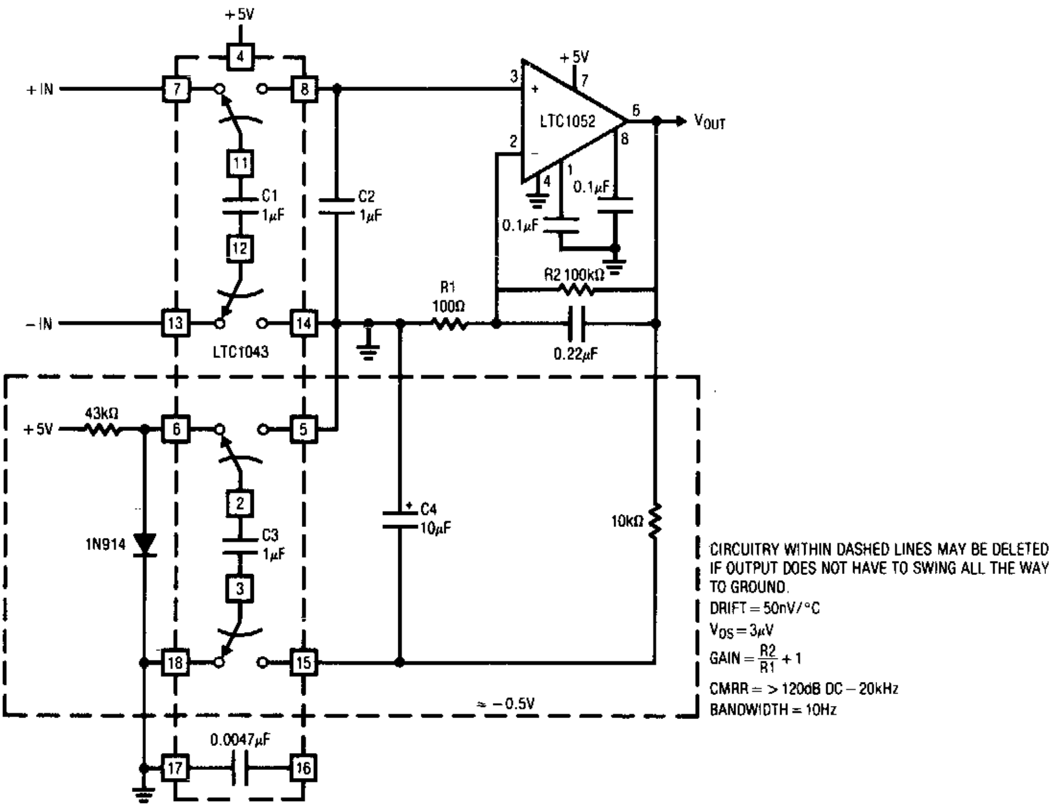
Fig. 45-8

46

Instrumentation Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

ULTRA-PRECISION INSTRUMENTATION AMPLIFIER

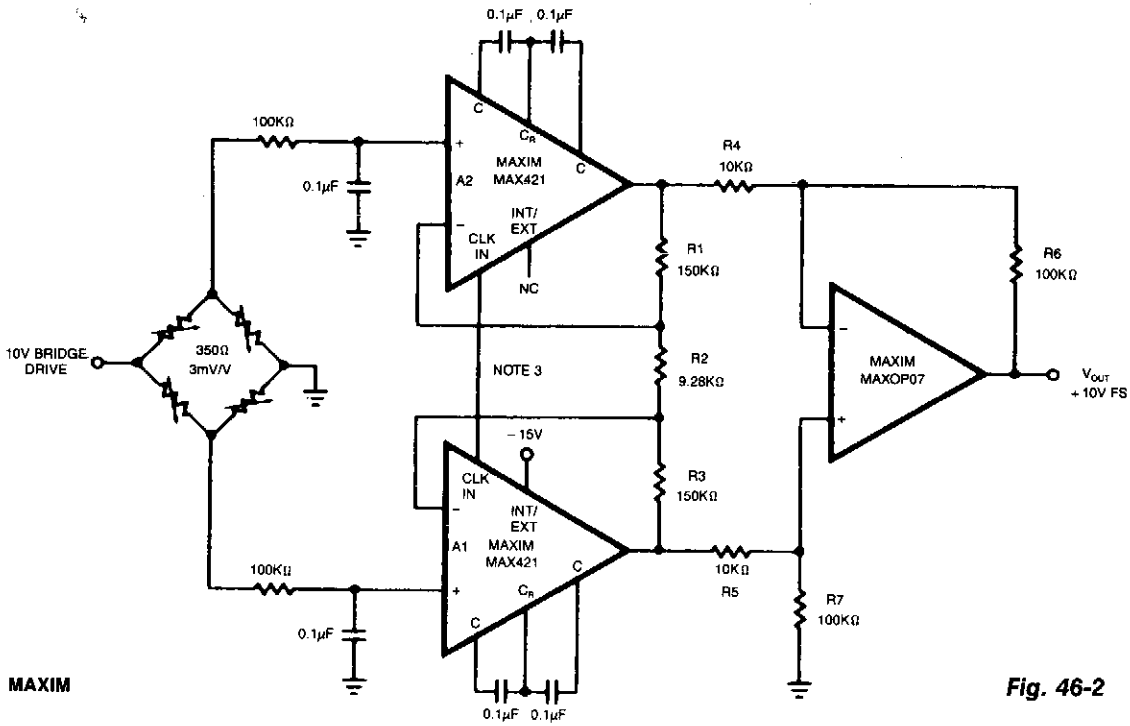


LINEAR TECHNOLOGY

Fig. 46-1

This circuit will run from a single 5 V power supply. The LTC1043 switched-capacitor instrumentation building block provides a differential-to-single-ended transition using a flying-capacitor technique. C1 alternately samples the differential input signal and charges ground referred C2 with this information. The LTC1052 measures the voltage across C2 and provides the circuit's output. Gain is set by the ratio of the amplifier's feedback resistors. Normally, the LTC1052's output stage can swing within 15 mV of ground. If operation all the way to zero is required, the circuit shown in dashed lines can be employed. This configuration uses the remaining LTC1043 section to generate a small negative voltage by inverting the diode drop. This potential drives the 10-KΩ, pull-down resistor, forcing the LTC1052's output into class A operation for voltages near zero. Note that the circuit's switched-capacitor front-end forms a sampled-data filter allowing the common-mode rejection ratio to remain high, even with increasing frequency. The 0.0047μF unit sets front-end switching frequency at a few hundred Hz.

STRAIN GAUGE INSTRUMENTATION AMPLIFIER

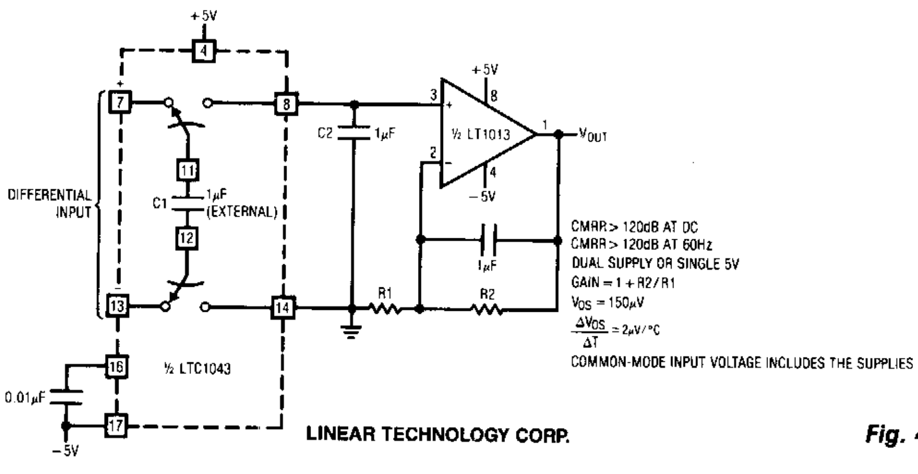


MAXIM

Fig. 46-2

This circuit has an overall gain of 320. More gain can easily be obtained by lowering the value of R2. Untrimmed V_{OS} is 10 μV, and V_{OS} tempco is less than 0.1 μV/°C. In many circuits, the OP07 can be omitted, with the two MAX421 differential outputs connected directly to the differential inputs of an integrating a/d.

INSTRUMENTATION AMPLIFIER

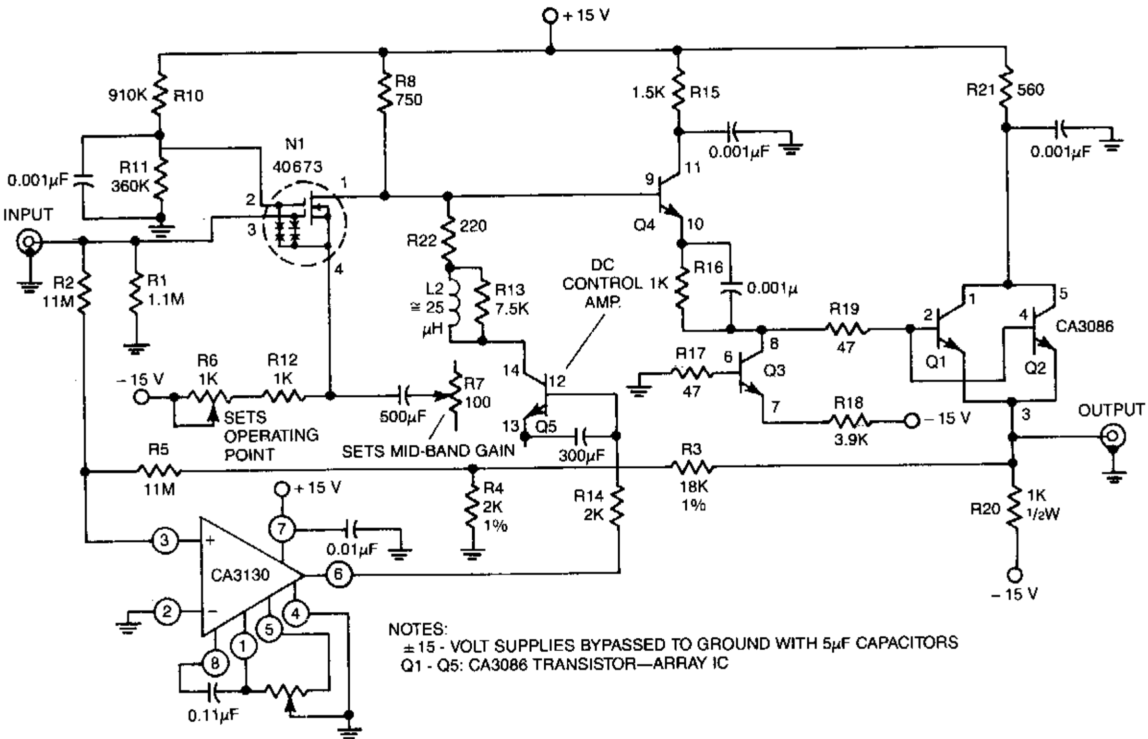


LINEAR TECHNOLOGY CORP.

Fig. 46-3

LTC1043 and LT1013 dual op amps are used to create a dual instrumentation amplifier using just two packages. A single DPDT section converts the differential input to a ground-referred single-ended signal at the LT1013's input. With the input switches closed, C1 acquires the input signal. When the input switches open, C2's switches close and C2 receives charge. Continuous clocking forces C2's voltage to equal the difference between the circuit's inputs. The 0.01-μF capacitor at pin 16 sets the switching frequency at 500 Hz. Common-mode voltages are rejected by over 120 dB and drift is low.

WIDEBAND INSTRUMENTATION AMPLIFIER

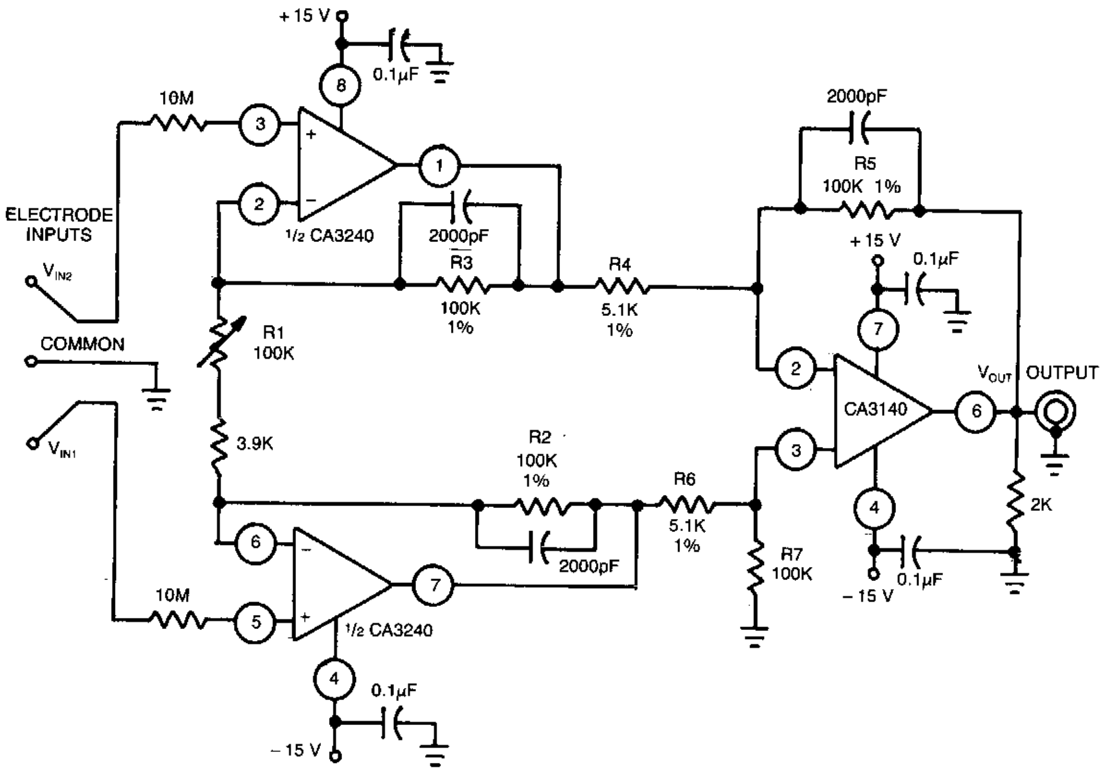


GE/RCA

Fig. 46-4

Has an input resistance of 1-MΩ, a bandwidth from dc to about 35 MHz, and a gain of 10 times. Low-frequency gain is provided by a CA3130 BiMOS op amp operated as a single-supply amplifier. High-frequency gain is provided by a 40673 dual-gate MOSFET. The entire amplifier is nulled by shorting the input to ground and adjusting R9 for zero dc output voltage.

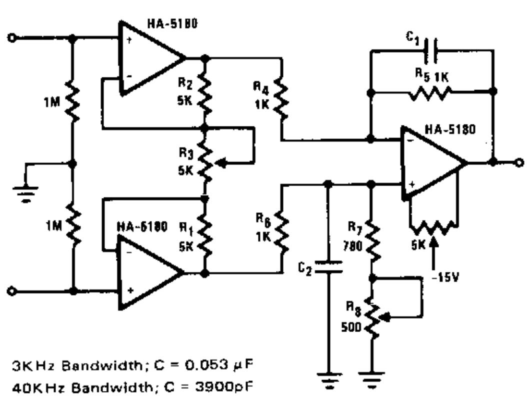
BIOMEDICAL INSTRUMENTATION DIFFERENTIAL AMPLIFIER



GE/RCA *Fig. 46-5*

This differential amplifier uses the isolated high-impedance inputs of the CA3420 BiMOS op amp. Because the CA3240's input current is only 50 pA maximum, 10-MΩ resistors can be used in series with the input probes to limit the current to 2 μA under a fault condition.

DIFFERENTIAL INSTRUMENTATION AMPLIFIER

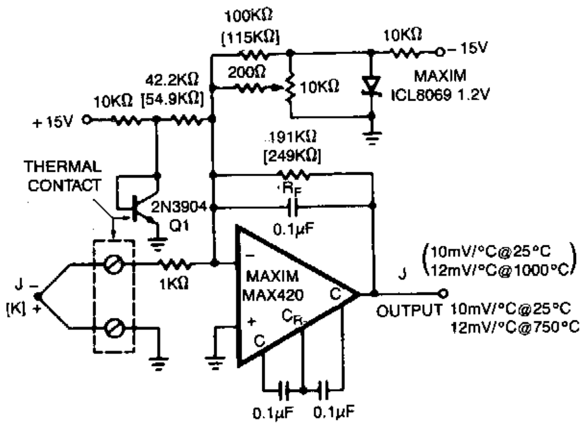


HARRIS

Fig. 46-6

This circuit relies on extremely high input impedance for effective operation. The HA-5180 with its JFET input stage, performs well as a preamplifier. The standard three amplifier configuration is used with very close matching of the resistor ratios *R*5/*R*4 and (*R*7 + *R*8)/*R*6, to insure high common-mode rejection (CMR). The gain is controlled through *R*3 and is equal to 2*R*1/*R*3. Additional gain can be had by increasing the ratios *R*5/*R*4 and (*R*7 + *R*8)/*R*6. The capacitors C1 and C2 improve the ac response by limiting the effects of transients and noise. Two suggested values are given for maximum transient suppression at frequencies of interest. Some of the faster DVM's are operating at peak sampling frequency of 3-kHz, hence the 4-kHz, low-pass time constant. The 40-kHz, low-pass time constant for ac voltage ranges is an arbitrary choice, but should be chosen to match the bandwidth of the other components in the system. C1 and C2 might however, reduce CMR for ac signals if not closely matched. Input impedances have also been added to provide adequate dc bias currents for the HA-5180 when open-circuited.

THERMOCOUPLE PREAMPLIFIER

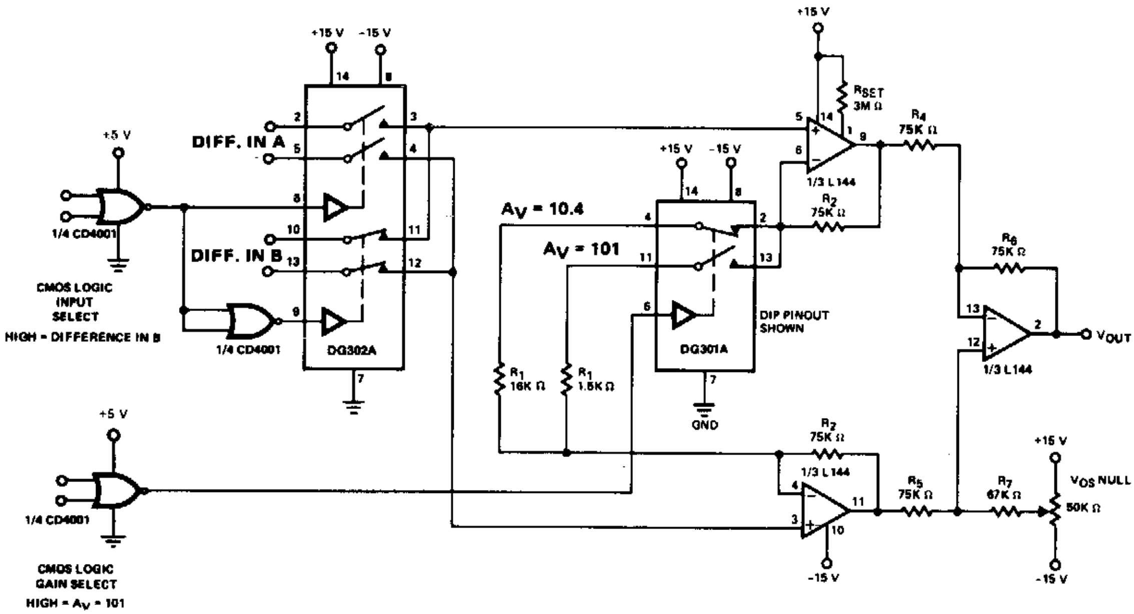


The MAX420 is operated at a gain of 191 to convert the 52 μV/°C output of the type J thermocouple to a 10 mV/°C signal. The −2.2 mV/°C tempco of the 2N3904 is added into the summing junction with a gain of 42.2 to provide cold-junction compensation. The ICL8069 is used to remove the offset caused by the 600-mV initial voltage of the 2N3904. Adjust the 10-KΩ trimpot for the proper reading with the 2N3904 and isothermal connection block at a temperature near the center of the circuit's operating range. Use the component values shown in parentheses when using a type K thermocouple.

MAXIM

Fig. 46-7

LOW-POWER INSTRUMENTATION AMPLIFIER



R_{SET} programs L144 power dissipation, gain-bandwidth product. Refer to AN73-6 and the L144 data sheet.

Voltage gain of the instrumentation amplifier is:

$$A_V = 1 + \frac{2R_2}{R_1} \quad \text{(In the circuit shown, } A_{V1} = 10.4,\ A_{V2} = 101\text{)}$$

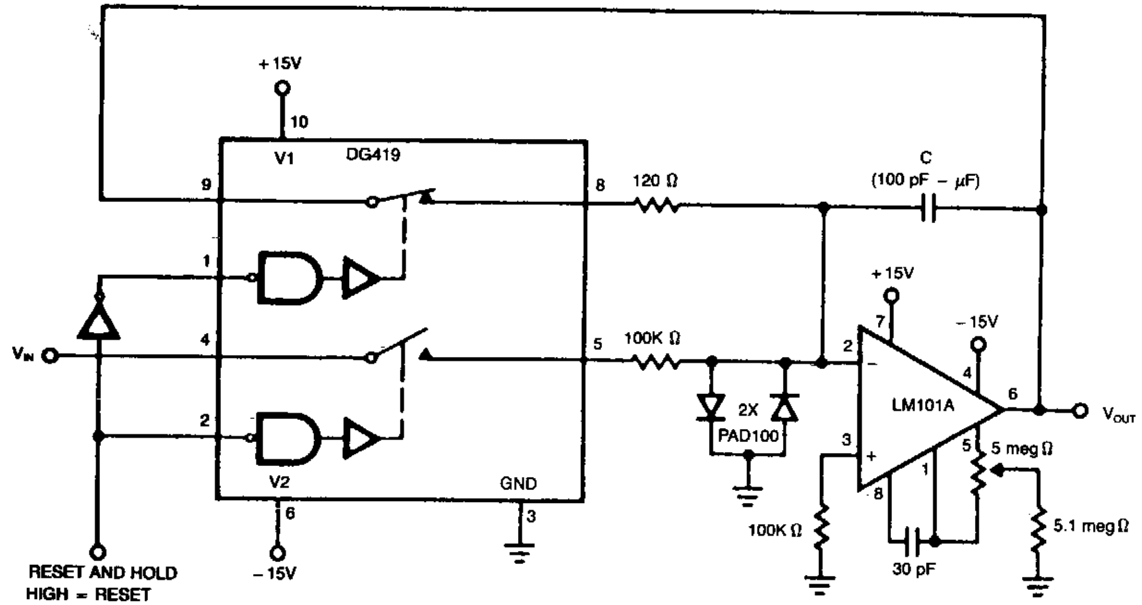
SILICONIX *Fig. 46-8*

47

Integrator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

RESETTABLE INTEGRATOR



NOTE: PIN CONNECTIONS SHOWN ARE FOR METAL CAN PACKAGE

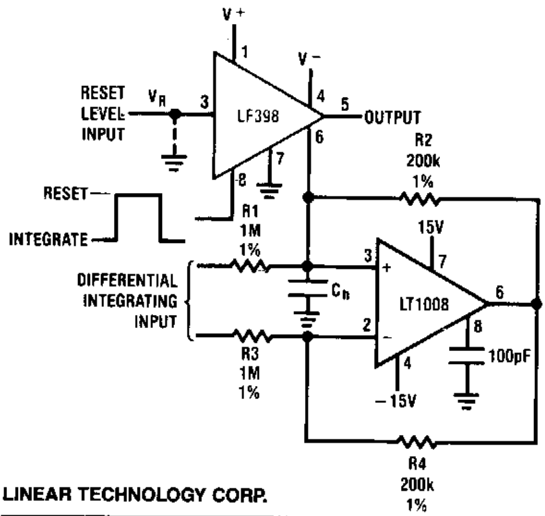
SILICONIX

$$V_O = \frac{10}{C}\int_{t_1}^{t_2} V_{IN}\,dt \qquad (C \text{ in } \mu F)$$

Fig. 47-1

The low $r_{DS(on)}$ and high peak current capability of the DG419 makes it ideal for discharging an integrator capacitor. A high logic input pulse disconnects the integrator from the analog input and discharges the capacitor. When the logic input lowers, the integrator is triggered. D1 and D2 prevent the capacitor from charging to over 15 V.

INTEGRATOR WITH PROGRAMMABLE RESET LEVEL



$$V_{OUT}\,(\text{HOLD MODE}) = \left[\frac{1}{(R1)C_h}\int_0^t V_{IN}\,dt\right] + \left[V_R\right]$$

LINEAR TECHNOLOGY CORP.

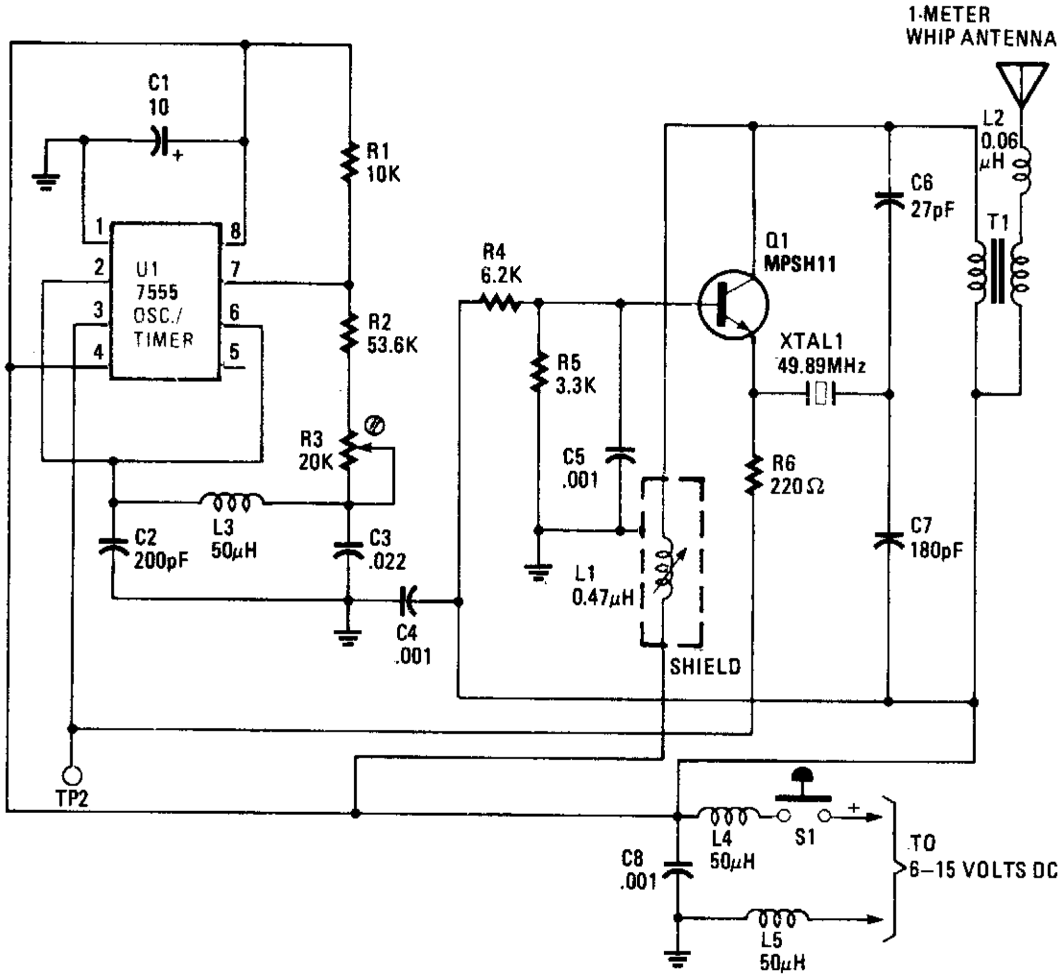
Fig. 47-2

48

Intercom Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

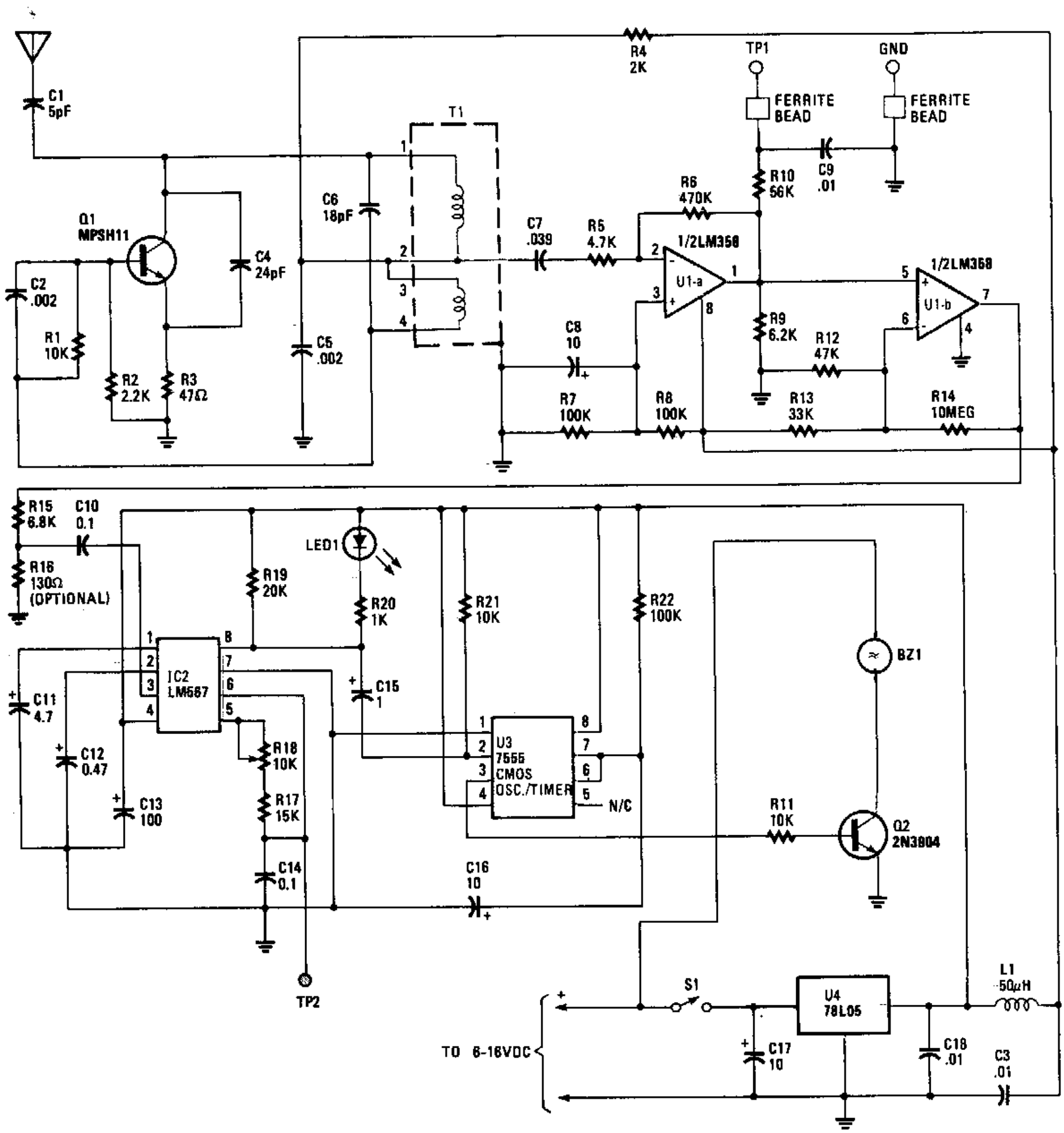
PERSONAL POCKET PAGER



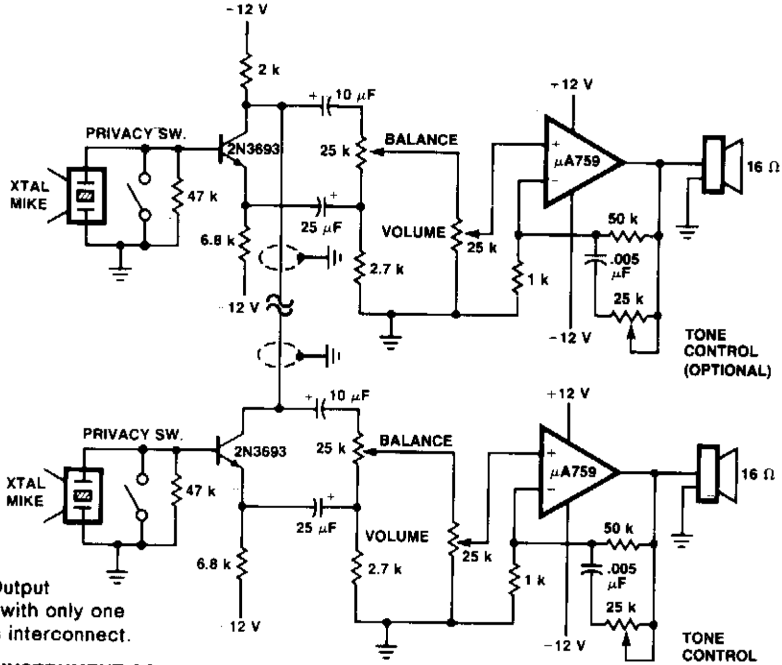
HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 48-1

When activated, the transmitter sends out a 49.890-MHz, AM rf carrier. The receiver detects, amplifies, and decodes the rf signal, which, in turn, activates a piezo buzzer. The receiver is small enough to carry in a pocket or sit on your workbench. The transmitter is also small and fits easily into a pocket for quick access.


C1
5pF
Q1
MPSH11
C2
.002
R1
10K
R2
2.2K
R3
47Ω
C4
24pF
C6
18pF
C5
.002
T1
R4
2K
TP1
GND
FERRITE BEAD
FERRITE BEAD
C9
.01
R10
56K
R6
470K
C7
.039
R5
4.7K
1/2LM358
U1-a
1/2LM358
U1-b
C8
10
R9
6.2K
R12
47K
R7
100K
R8
100K
R13
33K
R14
10MEG
R15
6.8K
C10
0.1
R16
130Ω
(OPTIONAL)
LED1
R19
20K
R20
1K
R21
10K
R22
100K
BZ1
IC2
LM567
C11
4.7
C12
0.47
C13
100
C15
1
R18
10K
R17
15K
C14
0.1
TP2
U3
7555
CMOS
OSC./TIMER
N/C
R11
10K
Q2
2N3904
C16
10
S1
U4
78L05
L1
50μH
TO 6-16VDC
C17
10
C18
.01
C3
.01

BIDIRECTIONAL INTERCOM SYSTEM



Features

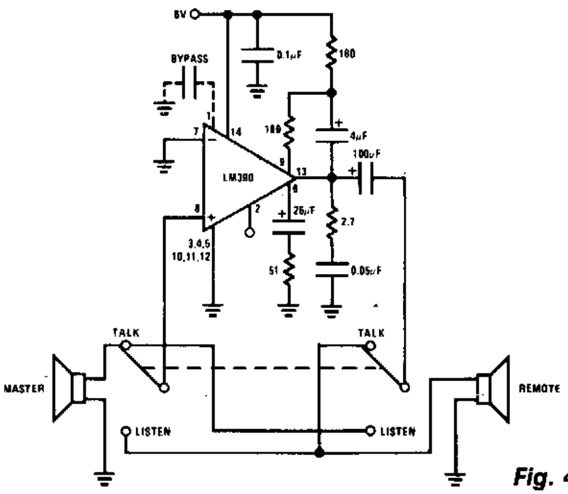
- Circuit Simplicity
- 1 Watt of Audio Output
- Duplex operation with only one two-wire cable as interconnect.

FAIRCHILD CAMERA & INSTRUMENT CORP.

Fig. 48-2

This system uses μA759 audio IC devices and a common connection between the preamps as an interconnect. Either mike can drive either speaker. Duplex operation is possible with only one cable (two wires).

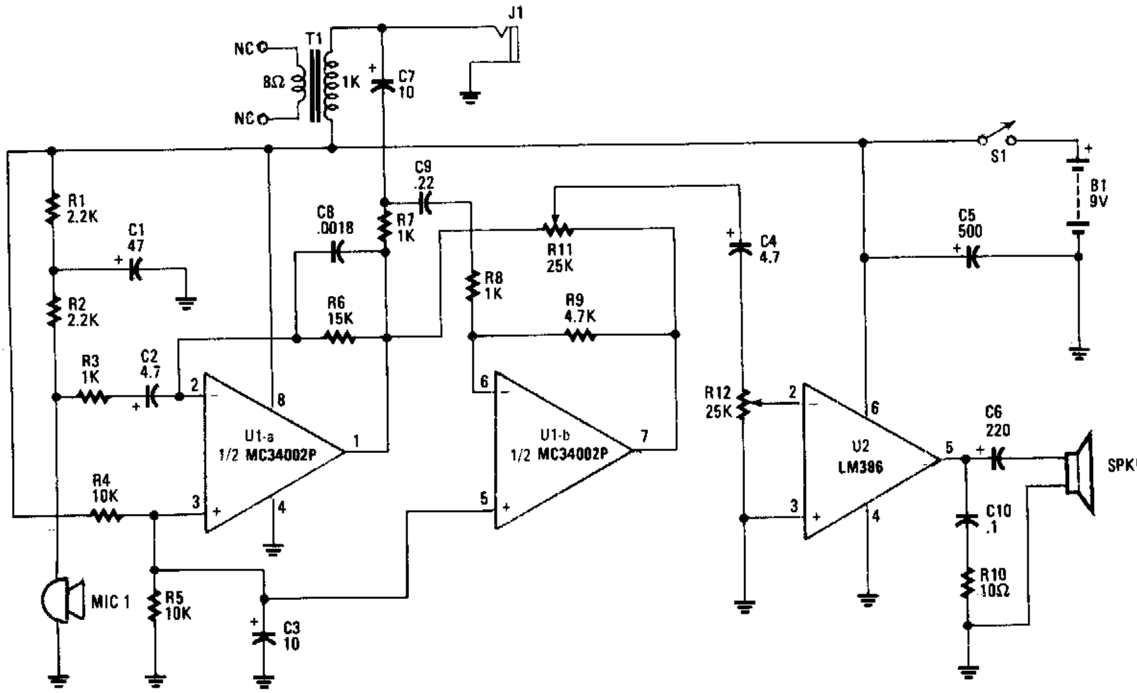
INTERCOM



This intercom uses a single audio IC as a two-way amplifier, and the speakers as microphones. A single 6-V supply provides adequate audio volume.

NATIONAL SEMICONDUCTOR CORP.

Fig. 48-3



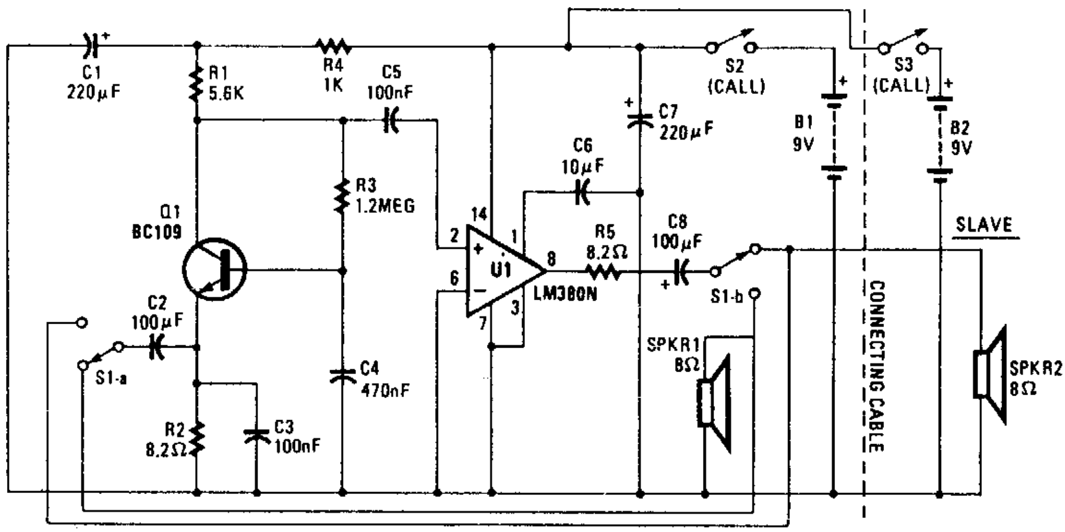
HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 48-4

Amplifier A increases the microphone's output to a usable level. The output signal is fed to op amp B, which inverts the signal 180°. A balance-control potentiometer connects across the outputs of amplifiers A and B. If an audio tone is fed into the microphone and the balance potentiometer's wiper is all the way over to the A output position, the tone will be heard at a high level. As the wiper is rotated toward the B output, the audio level will decrease until it just about disappears near the center of the potentiometer's range. As you continue to rotate the wiper, the signal will begin to increase once again.

With the balance control set for a minimal output, the intercom's tendency to self-oscillate from acoustical feedback between the microphone and speaker is kept to a minimum. The microphone's amplified signal at A's output is fed to the other intercom through the audio in/out cable. Since both intercom units are alike, the audio information coming from one unit feeds the other at the input of op amp B. The incoming audio is amplified slightly by op amp B and the output signal is sufficiently increased by the power amp to drive the speaker.

TWO-WAY INTERCOM



POPULAR ELECTRONICS

Fig. 48-5

49

Inverters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

12 VDC-TO-117 VAC AT 60 Hz POWER INVERTER

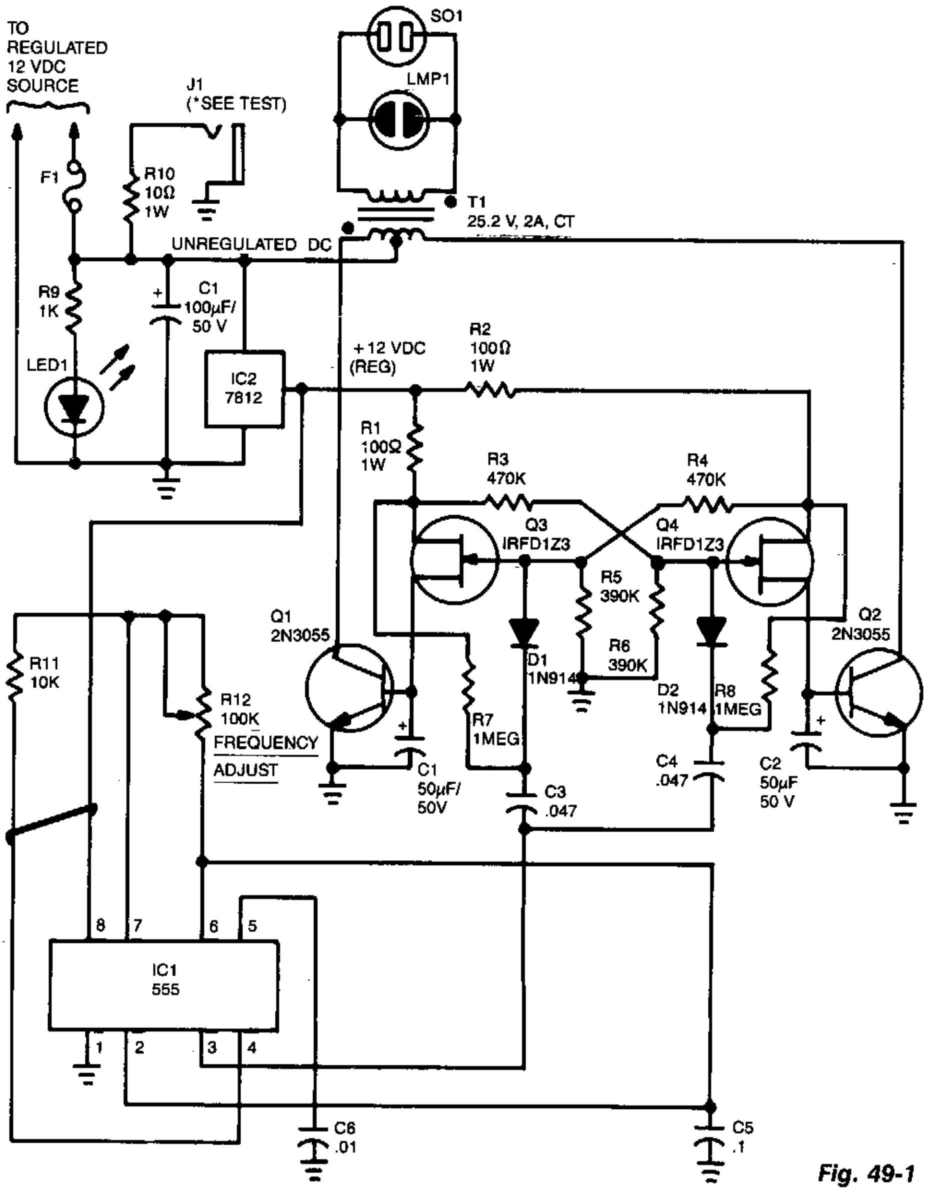


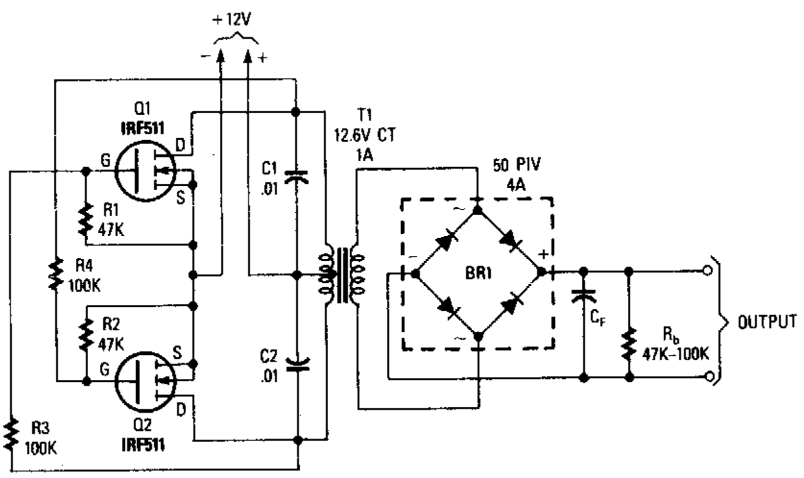
Fig. 49-1

Capacitor C5 and potentiometer R12 determine the frequency of the output signal at pin 3 of IC1, the 555 oscillator. The output signal is differentiated by C3 and C4 before it's input to the base of power transistors Q1 and Q2 via diodes D1 and D2, respectively. The signal from IC1 is adjusted to 120 Hz, because the flip-flop formed by transistors Q3 and Q4 divides the frequency by 2.

When Q3 is on, the base of Q1 is connected via R1 to the regulated 12-V supply. Then, when the flip-flop changes states, Q4 is turned on and the base of Q2 connected to the 12-V supply through R2. The 100 mA base current allows Q1 and Q2 to alternately conduct through their respective halves to the transformer's secondary winding.

To eliminate switching transients caused by the rapid switching of Q3 and Q4, capacitors C1 and C2 filter the inputs to the base of Q1 and Q2 respectively. Power for the unit comes from an automobile's 12-V system or from a storage battery. The power is regulated by IC2, a 7812 regulator. LED1, connected across the 12-V input, can be used to indicate whether power is being fed to the circuit. The neon pilot lamp, LMP1, shows a presence or absence of output power.

POWER MOSFET INVERTER

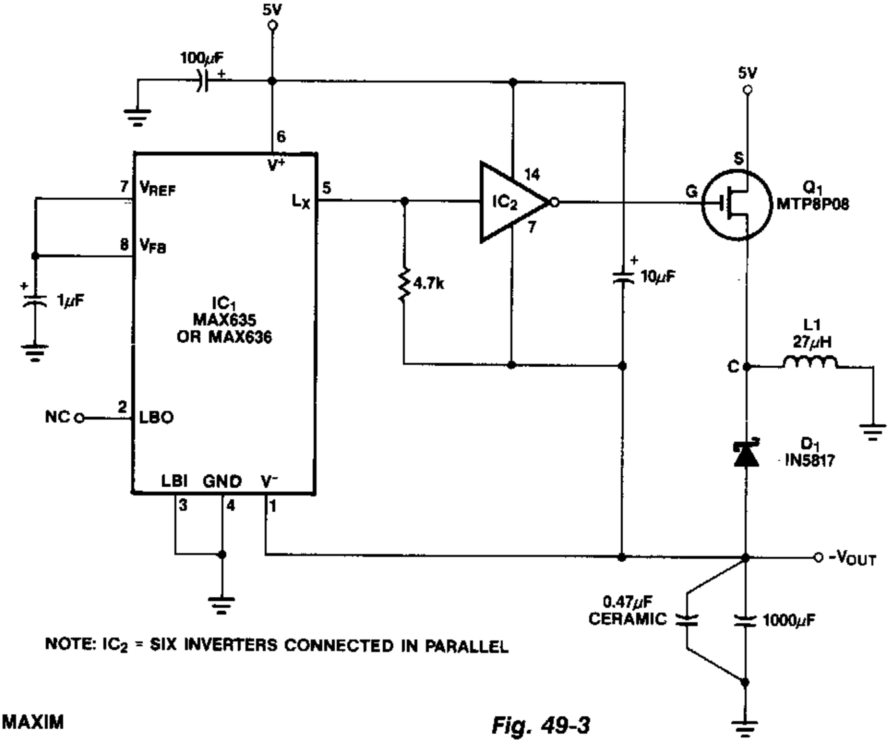


POPULAR ELECTRONICS

Fig. 49-2

This inverter can deliver high-voltage ac or dc, with a rectifier and filter, up to several hundred volts. The secondary and primary of T1—a 12.6 to 440 V power transformer, respectively—are reversed; e.g., the primary becomes the secondary and the secondary becomes the primary. Transistors Q1 and Q2 can be any power FET. Be sure to heat sink Q1 and Q2. Capacitors C1 and C2 are used as spike suppressors.

MEDIUM POWER INVERTER



NOTE: IC2 = SIX INVERTERS CONNECTED IN PARALLEL

MAXIM

Fig. 49-3

In this circuit, a CMOS inverter, such as the CD4069, is used to convert the open drain L_x output to a signal suitable for driving the gate of an external P MOSFET. The MTP8P03 has a gate threshold voltage of 2.0 V to 4.5 V, so it will have a relatively high resistance if driven with only 5 V of gate drive. To increase the gate drive voltage, and thereby increase efficiency and power handling capability, the negative supply pin of the CMOS inverter is connected to the negative output, rather than to ground. Once the circuit is started, the P MOSFET gate drive swings from +5 V to $-V_{OUT}$. At start up, the $-V_{OUT}$ is one Schottky diode drop above ground and the gate drive to the power MOSFET is slightly less than 5 V. The output should be only lightly loaded to ensure start up, since the output power capability of the circuit is very low until $-V_{OUT}$ is a couple of volts.

This circuit generates complementary output signals from 50 to 240 Hz. Digital timing control ensures a separation of 10 to 15° between the fall time of one output and the rise time of the complementary output.

The digital portion of inverter U1 to U4 controls the drive to Q1 and Q2, both MTE60N20 TMOS devices. These devices are turned on alternately with 11.25° separation between complementary outputs. A +12-V supply for CMOS gates U1 to U4 is developed by T1, D3, D4, C7, and U6. The power supply for the TMOS frequency generator is derived from the diode bridge, U5, and capacitor C7; it is applied to the center tap of T2.

COMPLEMENTARY OUTPUT VARIABLE FREQUENCY INVERTER

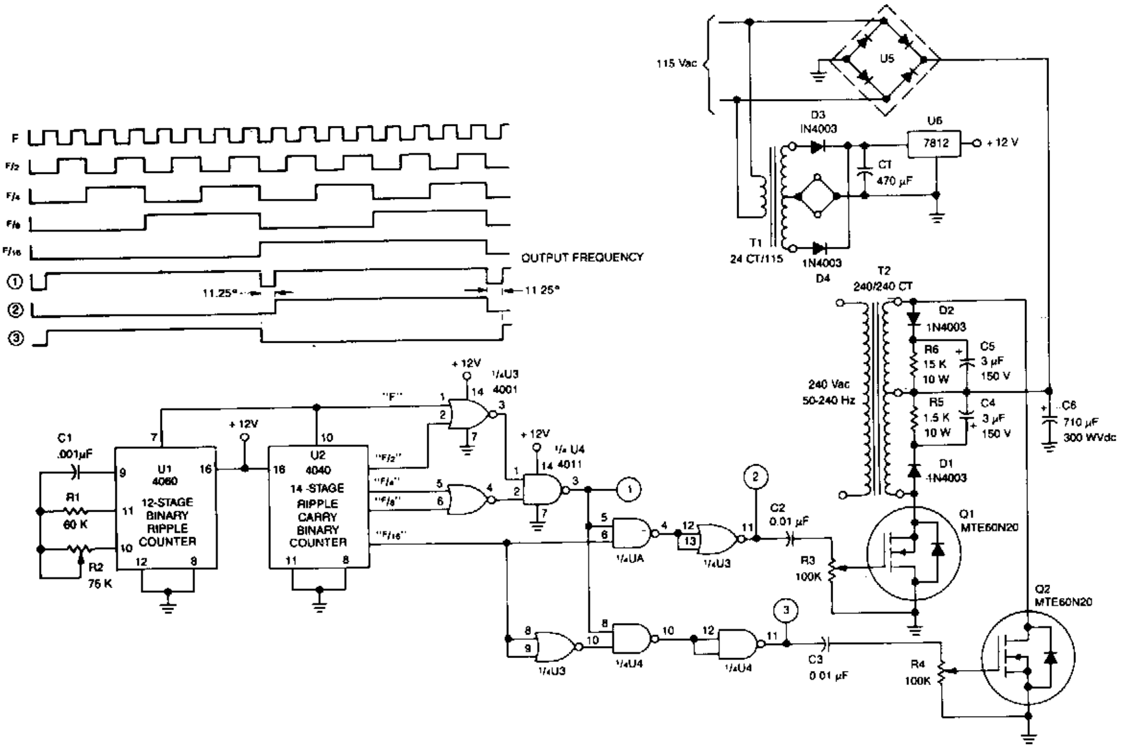
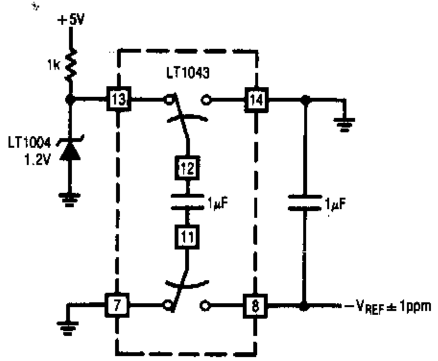


Fig. 49-4

U1 is a 4060 12-stage binary ripple counter that is used as a free-running oscillator; its frequency of oscillation is: 1/2.2 $C1R2$. The output of U1 is applied to U2, a 14-stage binary ripple counter that provides square-wave outputs of 1/2, 1/4, 1/8, and 1/16 of the clock frequency. These signals are combined in U3 and U4 to provide a complementary drive for Q1 and Q2.

Outputs from U3 and U4 are ac-coupled to Q1 and Q2 via C2 and C4, respectively. R3 and R4 adjust the gate drive to Q1 and Q2. Q1 and Q2 alternately draw current through opposing sides of the primary to synthesize an ac input voltage at a given frequency. Only one side of the primary of T2 is driven at one time, so maximum power output is half of the transformer rating.

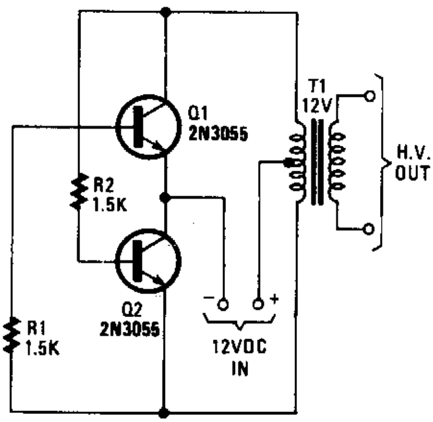
PRECISION VOLTAGE INVERTER



LINEAR TECHNOLOGY CORP. *Fig. 49-5*

This circuit allows a reference to be inverted with 1 ppm accuracy, features high input impedance, and requires no trimming.

POWER INVERTER



POPULAR ELECTRONICS *Fig. 49-6*

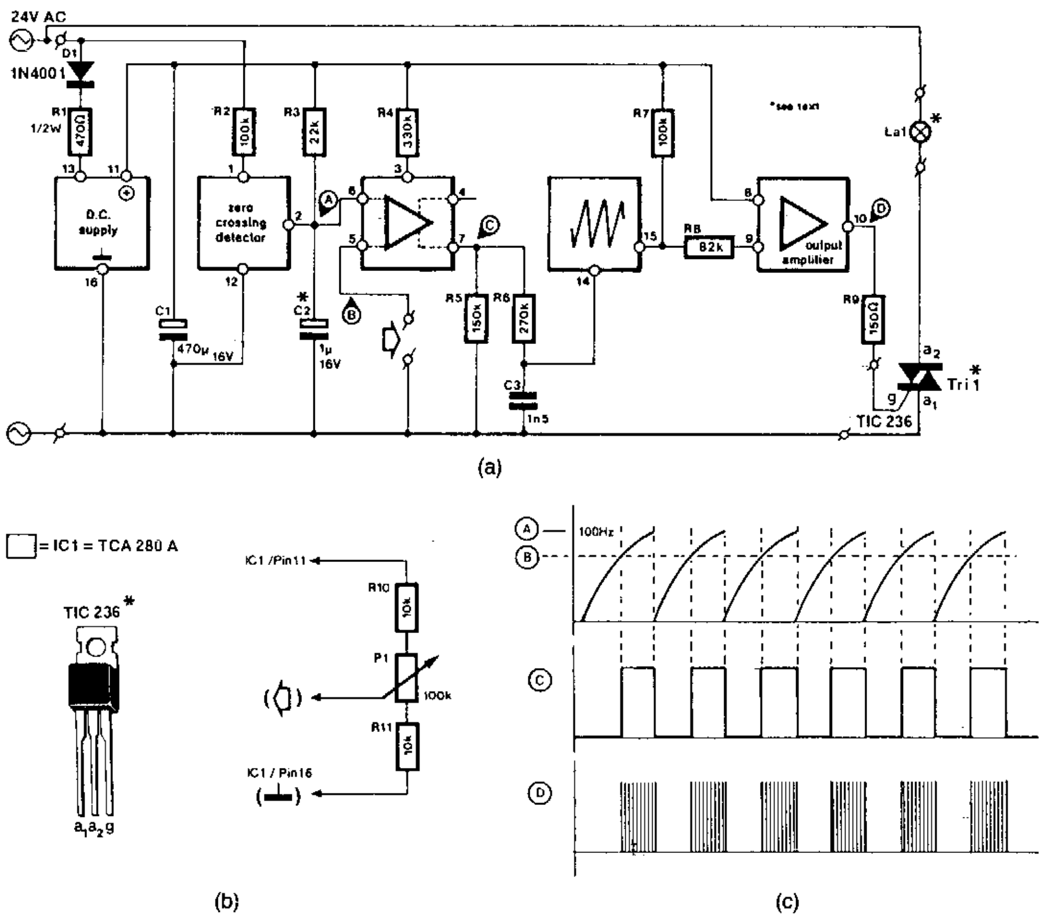
The transformer can be any 6.3 or 12.6 V type. Apply the 12-Vdc input so the positive goes to the transformer's center tap and the negative goes to the two transistor emitters. Any bridge-type rectifier and filter can be used at the output, if you need dc.

50

Lamp-Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

HALOGEN LAMP DIMMER

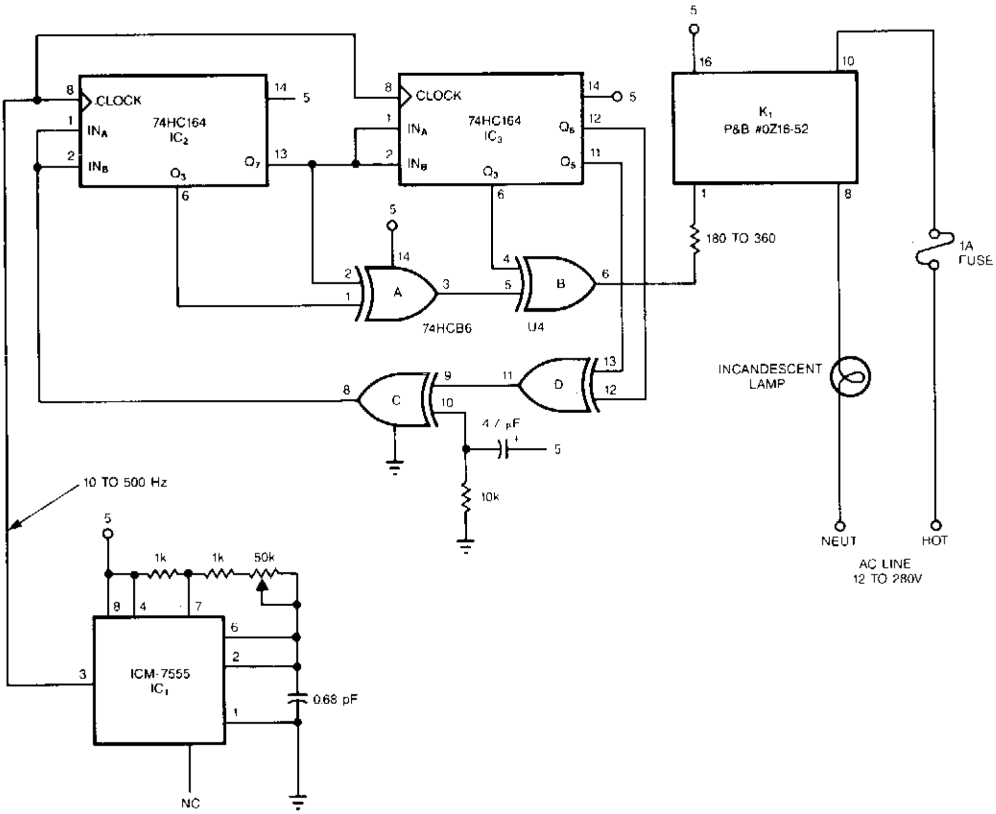


ELEKTOR ELECTRONICS *Fig. 50-1*

This circuit is suitable for fitting into slide projectors without a dimmer facility as with 24-Vac fed halogen lamps. With a few small alterations, it can also be used for dimming 12-V halogen lamps, but not those in a car, because these are fed from a dc source. The circuit shown in Fig. 50-1a is intended for operation from a 24-Vac supply, and can handle a lamp load of up to 150 W. For loads up to 250 W, the TIC236 should be replaced by a TIC246.

Figure 50-1b shows detail of the connection of a potentiometer to the intensity control input of the TCA280A. Voltage divider R10-P1-R11 is fitted externally and can be fed from the stabilized voltage available at pin 11 of IC1. The minimum and maximum intensities of the lamp are determined by R10 and R11, respectively.

PSEUDORANDOM SEQUENCER



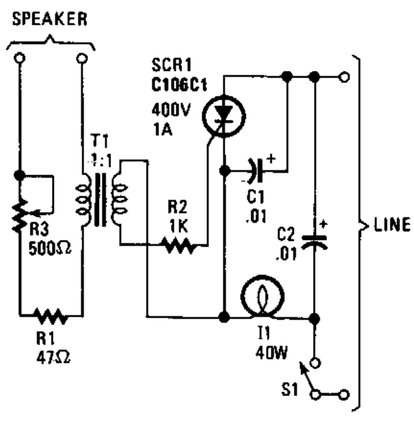
EDN

Fig. 50-2

This pseudorandom sequencer drives a solid-state relay. If you power a low-wattage lamp from the relay, the lamp will appear to flicker like a candle's flame in the wind; using higher-wattage lamps allows you to simulate the blaze of a fireplace or campfire. You can enhance the effect by using three or more such circuits to power an array of lamps.

The circuit is comprised of an oscillator IC1 and a 15-stage, pseudorandom sequencer, IC2 through IC4. The sequencer produces a serial bit stream that repeats only every 32, 767 bits. Feedback from the sequencer's stages 14 and 15 go through IC4D and back to the serial input of IC2. Note the rc network feeding IC4C; the network feeds a positive pulse into the sequencer to ensure that it won't get stuck with all zeros at power-up. The leftover XOR gates IC4A and IC4B further scramble the pattern. The serial stream from IC4B drives a solid-state relay that features zero-voltage switching and can handle loads as high as 1 A at 12 to 280 Vac.

LIGHT MODULATOR



HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 50-3

The lights seem to dance in time with the music. Line-voltage lamps of about 40 to 100 W do nicely. The current for the lamp is from an SCR. When low-level audio is present across T1, SCR1 is not triggered into conduction. A louder signal, however, triggers the SCR so that the lamp lights and follows the sounds. Since SCR1 is operated by an alternating current, the rectifier moves out of the avalanche condition when the gate current is low. Potentiometer R3 lets you adjust the power reaching transformer T1, so that with normal operating volume, SCR1 triggers again and again, except during quiet passages.

LAMP LIFE EXTENDER

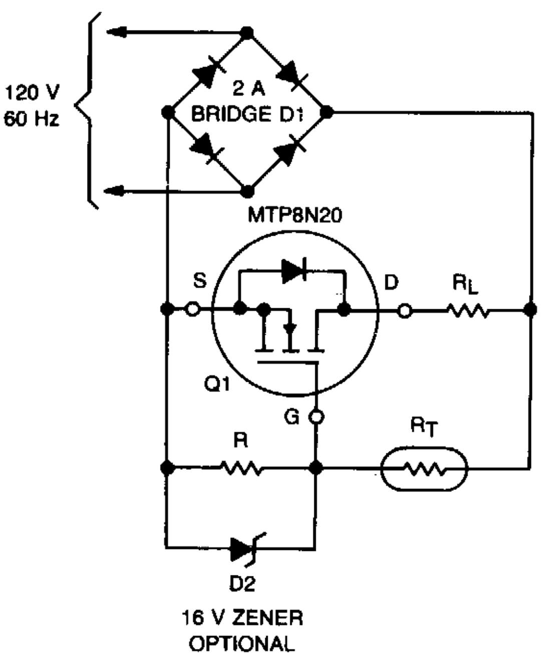


Fig. 50-4

Lamp life can be extended by improving the conditions under which its filament is operated. This includes eliminating the inrush overcurrent surge and reducing the mechanical stress (vibration) on the filament caused by an ac source.

The circuit shown controls the inrush current to the lamp without the 10 to 15 times-rated current stage that normally occurs when power is applied to a cold lamp. It does so by adjusting the inrush current over time to the inverse of the value normally experienced.

R_L is a standard tungsten lamp in the range of 15 to 250 W, R is 10-Ω and R_T is a negative temperature coefficient resistance that is initially 1.65 MΩ and decreases, by self-heating, to 150 KΩ in approximately 0.5 s. Use of the TMOS device allows high Ω values for R and R_T, keeping drive power at a negligible level.

This circuit has a number of advantages: very low power dissipation, long life, low-cost components, no significant effect on lamp ratings, negligible effect on efficiency, negligible RFI, and it can be used in hazardous environments.

PHASE CONTROL

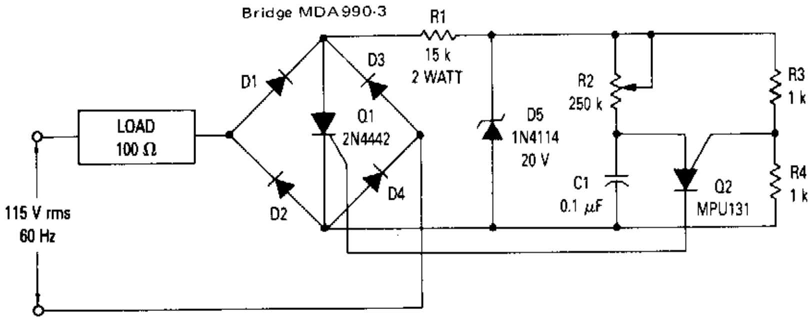
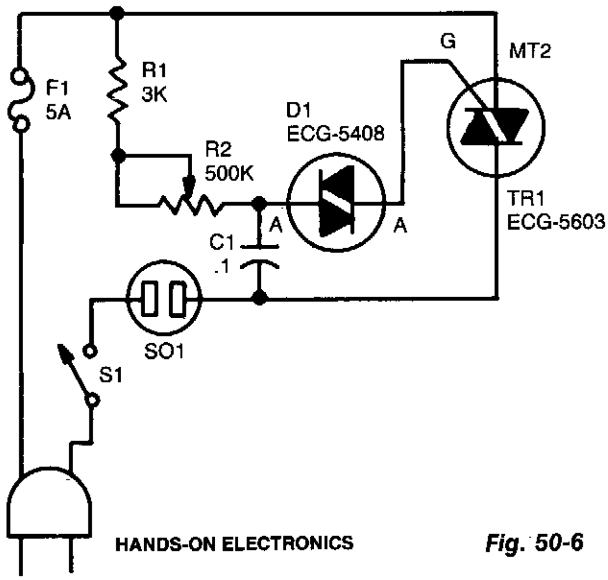


Fig. 50-5

This circuit uses a PUT for phases control of an SCR. The relaxation oscillator formed by Q2 provides conduction control of Q1 from 1 to 7.8 ms or 21.6° to 168.5°. This constitutes control of over 97% of the power available to the load. Only one SCR is needed to provide phase control for both the positive and negative portion of the sine wave by putting the SCR across the bridge—composed of diodes D1 through D4.

TRIAC LIGHT DIMMER



HANDS-ON ELECTRONICS

Fig. 50-6

800-W SOFT-START LIGHT DIMMER

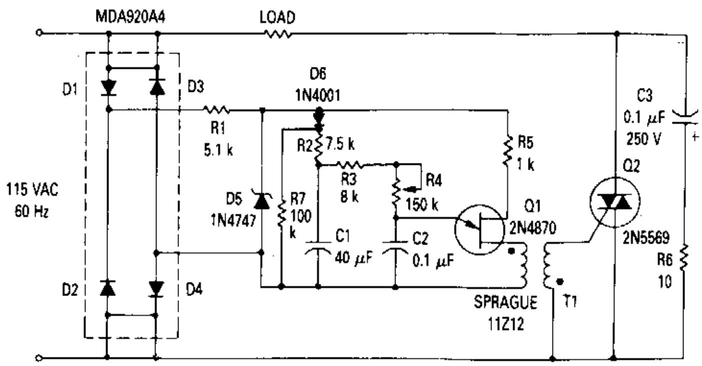
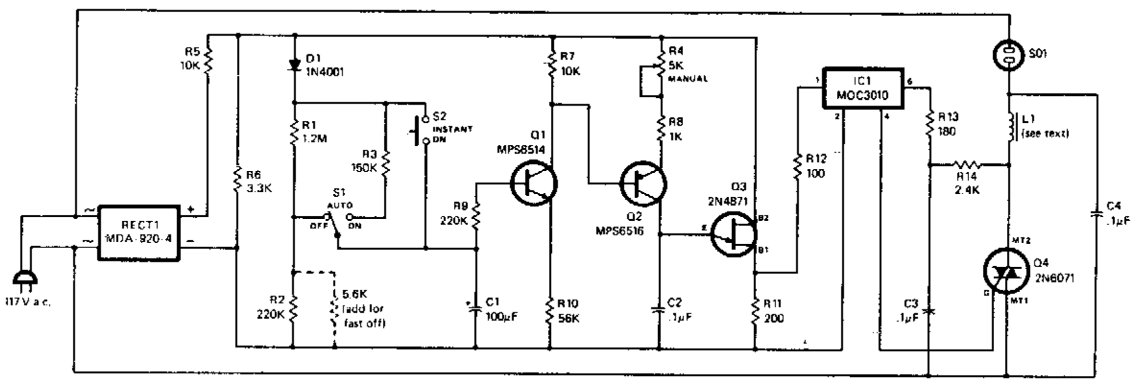


Fig. 50-7

The zener provides a constant voltage of 20 V to unijunction transistor Q1, except at the end of each half-cycle of the input when the line voltage drops to zero. Initially, the voltage across capacitor C1 is zero and capacitor C2 cannot charge to trigger Q1. C1 will begin to charge, but because the voltage is low, C2 will be charged to a voltage adequate to trigger C1 only near the end of the half cycle. Although the lamp resistance is low at this time, the voltage applied to the lamp is low and the inrush current is small. Then the voltage on C1 rises, allowing C2 to trigger Q1 earlier in the cycle. At the same time, the lamp is being heated by the slowly increasing applied voltage. By the time the peak voltage applied to the lamp has reached its maximum value, the bulb has been heated sufficiently to keep the peak inrush current at a reasonable value. Resistor R4 controls the charging rate of C2 and provides the means to dim the lamp. Diode D6 and resistor R7 improve operation at low-conduction angles.

SOLID-STATE LIGHT DISSOLVER

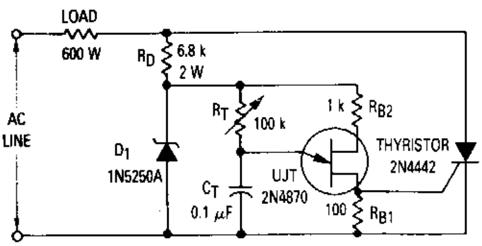


MODERN ELECTRONICS

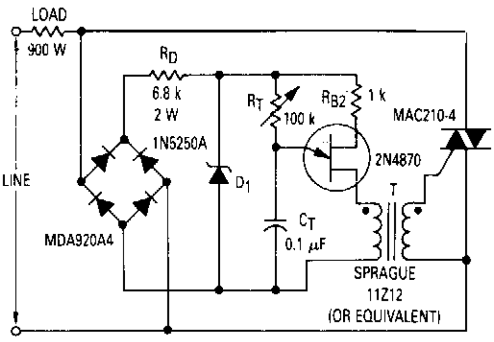
Fig. 50-8

The dimming action is controlled by varying the amount of current passed through triac Q4 and, thus, the lamp plugged into ac receptacle SO1. Unijunction transistor Q3 operates as a relaxation oscillator whose output pulse frequency depends on how fast capacitor C2 recharges after firing. Transistors Q1 and Q2 furnish the charging current, with the R3/C1 and R1/R2/C1 time-constant networks controlling the turn-on and turn-off times. Inside IC1 is a LED, a detector, and a small triac. In circuit, the low-level pulses coming from Q3 make the LED in IC1 emit short bursts of light that are picked up and converted into electrical current pulses by the internal detector. This small current triggers the internal triac, which then outputs the pulses to the gate of power triac Q4, triggering it on so that it delivers current to the lamp. Potentiometer R4 serves as a master control of the pulse rate and provides both manual control and a limit in the brightness of the lamp plugged into SO1. Momentarily pressing S2 causes the lamp to instantly turn on. Choke L1 suppresses any spikes produced by the power triac and limits interference with AM radio reception. No safeguards against interference need to be made for FM and TV reception, since these media are immune to this type of noise.

PHASE CONTROL



(a)



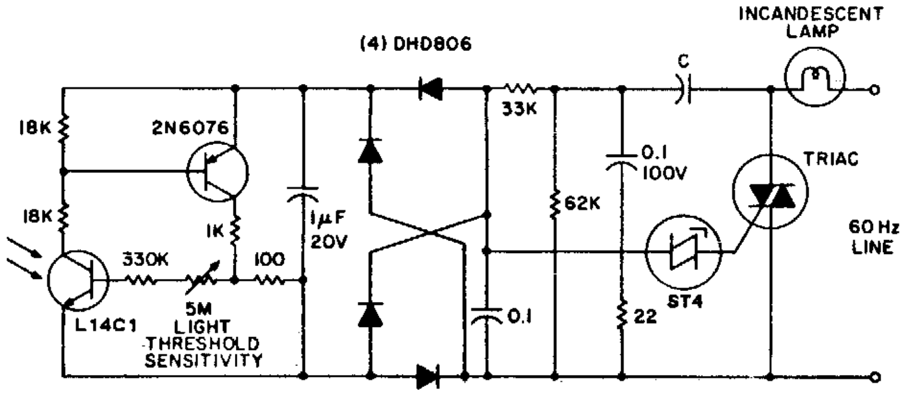
(b)

Fig. 50-9

The most elementary application is a half-wave control circuit. The thyristor is acting both as a power control device and as a rectifier, providing variable power to the load during the positive half cycle and no power to the load during the negative half cycle. The circuit is designed to be a two terminal control which can be inserted in place of a switch. If full-wave power is desired as the upper extreme of this control, a switch can be added which will short circuit the SCR when R_T is turned to its maximum power position. Full-wave control might be realized by the addition of a bridge rectifier, a pulse transformer, and by changing the thyristor from an SCR to a TRIAC, shown in Fig 50-9b. In this circuit, R_{B1} is not necessary since the pulse transformer isolates the thyristor gate from the steady-state UJT current. Occasionally, a circuit is required to provide constant output voltage regardless of line voltage changes. Adding potentiometer P1 to the circuits will provide an approximate solution to this problem. The potentiometer is adjusted to provide reasonably constant output over the desired range of line voltage.

LINE-VOLTAGE OPERATED AUTOMATIC NIGHT LIGHT

SUGGESTED TRIAC	LAMP WATTAGE MAXIMUM	
	120V	220V
SC141D	400W	800W
SC146D	550W	1100W
SC151D	750W	1500W
SC260D	1200W	2500W
SC265D	2000W	4000W

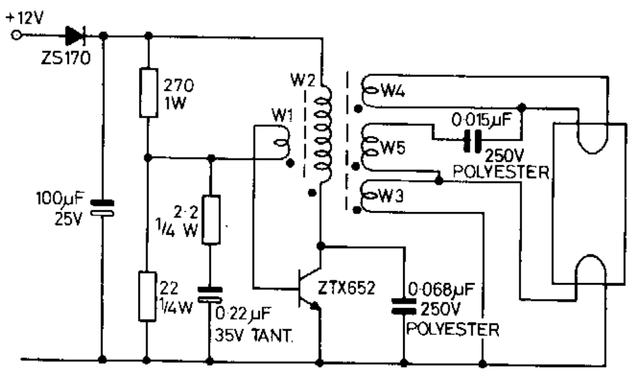


GE

Fig. 50-10

This circuit has stable threshold characteristics from its dependence on the photo diode current in the L14C1 to generate a base emitter voltage drop across the sensitivity setting resistor. The double phase shift network supplying voltage to the ST-4 trigger insures triac triggering at line voltage phase angles small enough to minimize RFI problems with a lamp load. This eliminates the need for a large, expensive inductor, contains the dV/dt snubber network, and utilizes lower voltage capacitors than the snubber of rfi suppression network normally used. The addition of a programmable unijunction timer can modify this circuit to turn the lamp on for a fixed time interval each time its environment gets dark. Only the additions to the previous circuit are shown in the interest of simplicity. When power is applied to the lamp, the 2N6028 timer starts. Upon completion of the time interval, the H11C3 is triggered and turns off the lamp by preventing the ST-4 from triggering the triac. The SCR of the H11C3 will stay on until the L14C1 is illuminated and allows the 2N6076 to commutate it off. Because of capacitor leakage currents, temperature variations and component tolerances, the time delay may vary considerably from nominal values.

8-W FLUORESCENT LAMP INVERTER

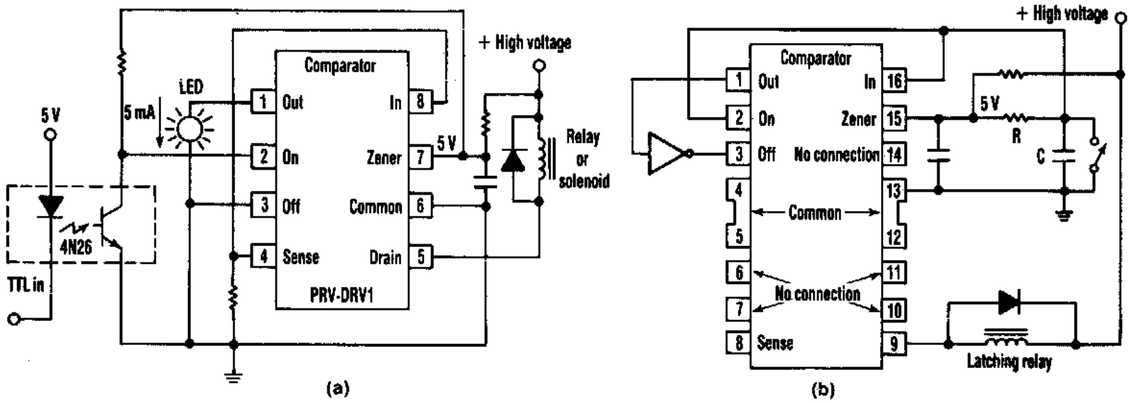


ZETEX, formerly FERRANTI

Fig. 50-11

This circuit has been designed to drive an 8-W fluorescent lamp from a 12-V source using an inexpensive inverter based on the ZTX652 transistor. The inverter will operate from supplies in the range of 10 to 16.5 V, thus making it suitable for use in on-charge systems such as caravanettes as well as periodically charged systems, such as camping lights, outhouse lights, etc. Other features of the inverter are an inaudible 20-kHz oscillator and reverse polarity protection.

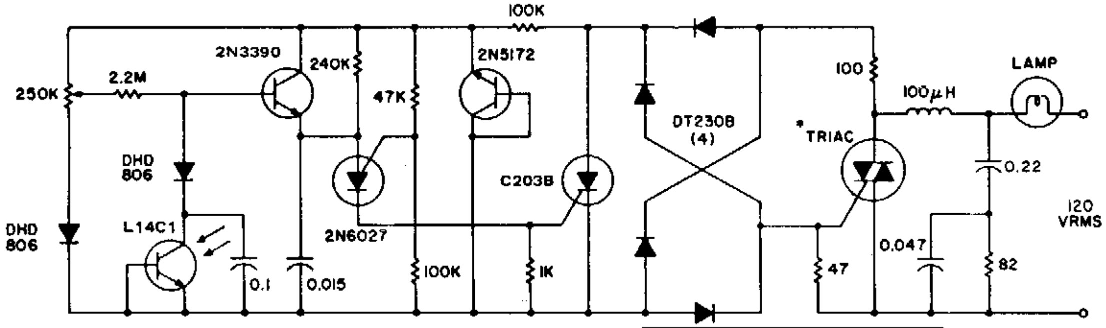
PULSE-WIDTH MODULATION LAMP BRIGHTNESS CONTROLLER



ELECTRONIC DESIGN *Fig. 50-12*

At half brightness, the lamp current is pulsed on and off (Fig. 50-12b) by the voltage developed across the resistor and capacitor at the current-sense output. Lamp current is sensed by the current-sense output. A simple pulse-width modulation lamp-brightness control circuit can also be built with the device. When the device powers up, the sense output is low, pulling the comparator output and the on input low, and turning the FET switch on. When the switch is on, current from the sense output charges the capacitor in the rc timing network to the 200-mV comparator threshold voltage. The comparator trips, turning the switch off. The charge then leaks off the capacitor, its voltage drops below 100 mV, and the FET is again turned on. The average current through the load is basically a function of the resistor value. The pulse-width modulation frequency on the other hand, is a function of the capacitor value.

CONSTANT BRIGHTNESS CONTROL



*The triac is matched to the lamp per chart

SUGGESTED TRIAC	LAMP WATTAGE MAXIMUM	
	120V	220V
SC141D	400W	800W
SC146D	550W	1100W
SC151D	750W	1500W
SC260D	1200W	2500W
SC265D	2000W	4000W

GE

Fig. 50-13

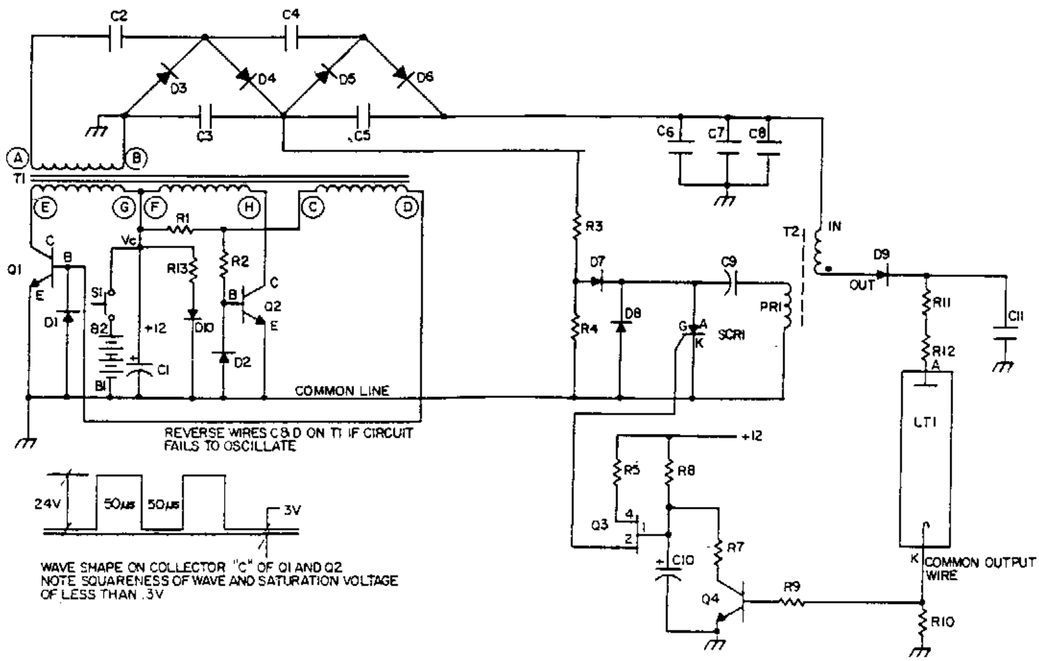
An automatic control maintains a lamp at a constant brightness over a wide range of supply voltages. This circuit utilizes the consistency of photodiode response to control the phase angle of power line voltage applied to the lamp and can vary the power between that available and ≈ 30% of available. This provides a candlepower range from 100% to less than 10% of nominal lamp output. The 100-μH choke, resistor, and capacitors form an rlc filter network that is used to eliminate conducted RFI.

51

Laser Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

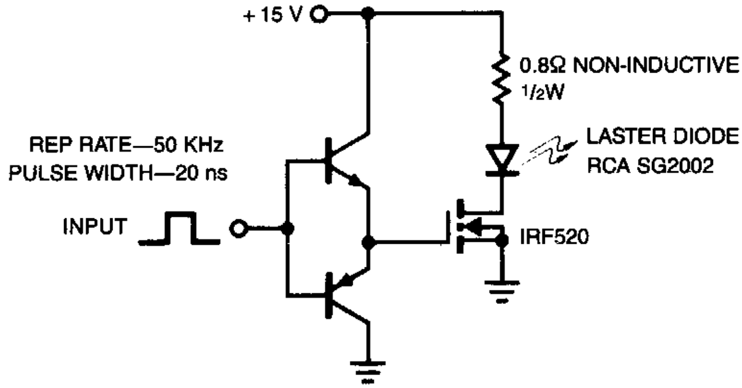
VISIBLE RED CONTINUOUS LASER GUN

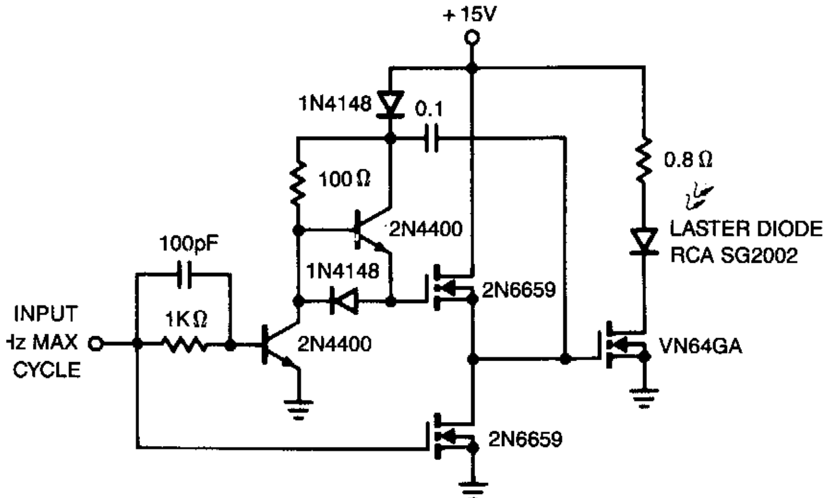


TAB BOOKS

Fig. 51-1

Q1 and Q2 switch the primary windings of transformer T1 via a square wave at a frequency determined by its magnetic properties. Diodes D1 and D2 provide base return paths for the feedback current of Q1 and Q2. The output winding of T1 is connected to a multiple section voltage multiplier. That multiplier consists of capacitors C1 through C5 and diodes D3 through D6. Resistors R3 and R4 divide the 800 V taken off at the junction of C3, and C5 for charging the dump capacitor C9 in the ignitor circuit. The ignitor, consisting of the T2 pulse transformer and capacitor discharge circuitry, provides the high-voltage dc pulse to ignite the laser the SCR1 dumping the energy of capacitor C9 into the primary of T2. The high-voltage pulses in capacitor C11 through rectifier diode D9. When C11 is charged to a LT1, ignition takes place and a current now flows that is sufficient to sustain itself at the lower voltage output of the voltage multiplier section. The path for this sustaining current is through the secondary of T2 and ballast resistors R11 and R12. The ignitor circuit is now deactivated by the clamping of Q3 emitter via Q4 being turned on by the voltage drop occurring across R10. This voltage drop will only occur when the laser tube is ignited and causes the SCR1 to cease firing; otherwise, the ignitor circuit would continue to operate, unnecessarily drawing on the limited power available.





SILICONIX

Fig. 51-2

The laser diode pulser is a simple drive circuit capable of driving the laser diode with 10-A, 20-ns pulses. For a 0.1% duty cycle, the repetition rate will be 50 kHz. A complementary emitter follower is used as a driver. Switching speed is determined by the f_T of the bipolar transistors used and the impedance of the drive source. A faster drive circuit is shown. It can supply higher peak gate current to switch the IRF520 very quickly. This circuit uses a MOSPOWER totem-pole stage to drive the high power switch. The upper MOSFET is driven by a bootstrap circuit. Typical switching times for this circuit are about 10 ns for both turn-on and turn-off.

52

Light-Controlled Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

FLAME MONITOR

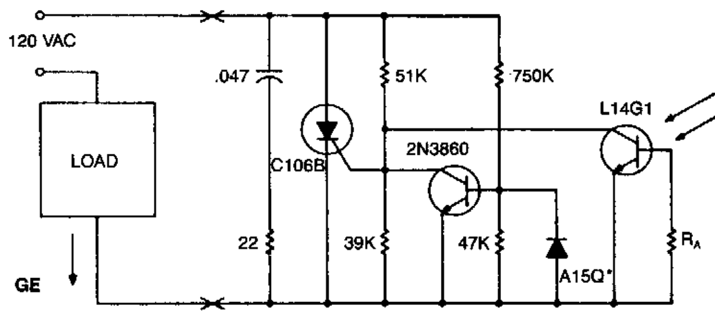
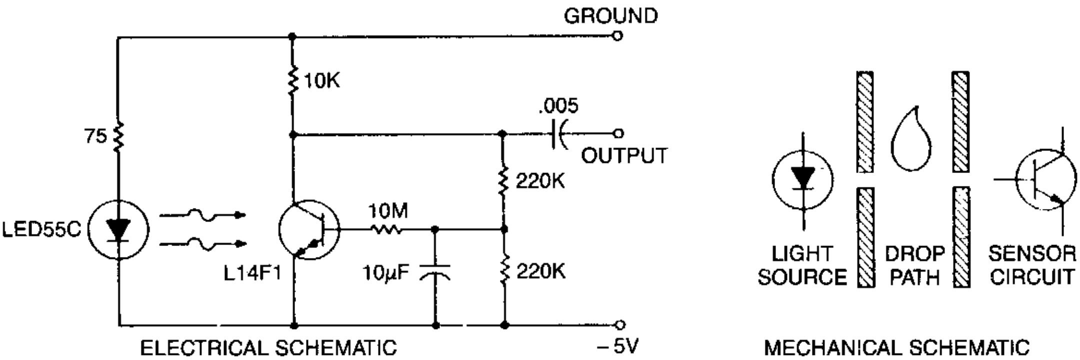


Fig. 52-1

Monitoring a flame and directly switching a 120-V load is easily accomplished using the L14G1 for point sources of light. For light sources which subtend over 10° of arc, the L14C1 should be used and the illumination levels raised by a factor of 5. This circuit provides zero voltage switching to eliminate phase controlling.

LOW-LIGHT LEVEL DROP DETECTOR



GE

Fig. 52-2

This self-biasing configuration is useful any time small changes in light level must be detected, for example, when monitoring very low flow rates by counting drops of fluid. In this bias method, the photodarlington is dc bias stabilized by feedback from the collector, compensating for different photodarlington gains and LED outputs. The 10-μF capacitor integrates the collector voltage feedback, and the 10-MΩ resistor provides a high base-source impedance to minimize effects on optical performance. The detector drop causes a momentary decrease in light reaching the chip, which causes collector voltage to momentarily rise, generating an output signal. The initial light bias is small because of output power constraints on the LED and mechanical spacing system constraints. The change in light level is a fraction of this initial bias because of stray light paths and drop translucence. The high sensitivity of the photodarlington allows acceptable output signal levels when biased in this manner. This compares with unacceptable signal levels and bias point stability when biased conventionally, i.e., base open and signal output across the collector bias resistor.

LIGHT-CONTROLLED LAMP SWITCH

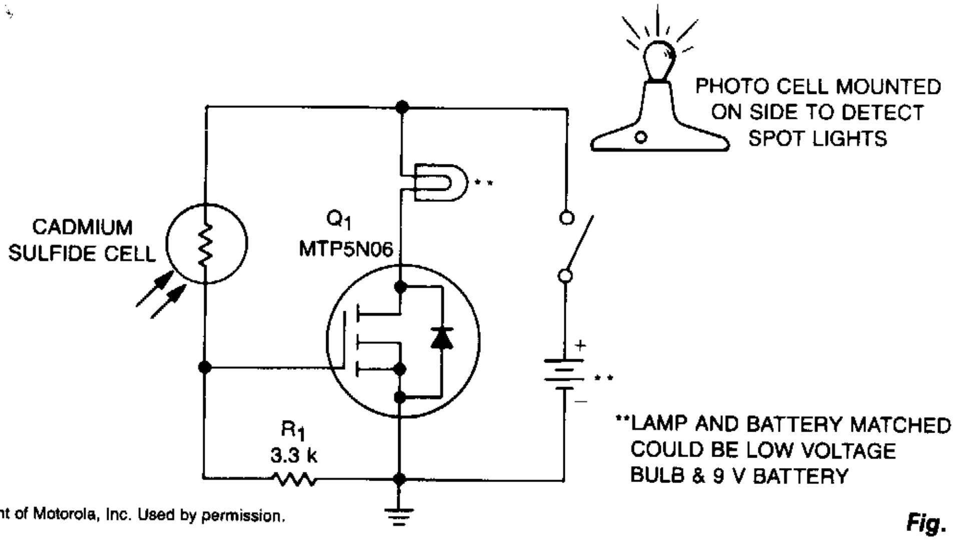


Fig. 52-3

A school drama needed lamps that automatically turned on and off when spot lights did the same. Lamp switching had to be wireless, durable, dependable, simple and inexpensive.

With stage and spot lights off, very little light falls on the CdS photocell, so its internal resistance is several megohms and R1 keeps the gate of Q1 at nearly zero volts, which keeps it off. When a spot or stage light hits the photocell, its resistance drops to several hundred ohms, raising Q1's gate voltage, which turns it on and applies power to the lamp.

OPTICAL SENSOR-TO-TTL INTERFACE

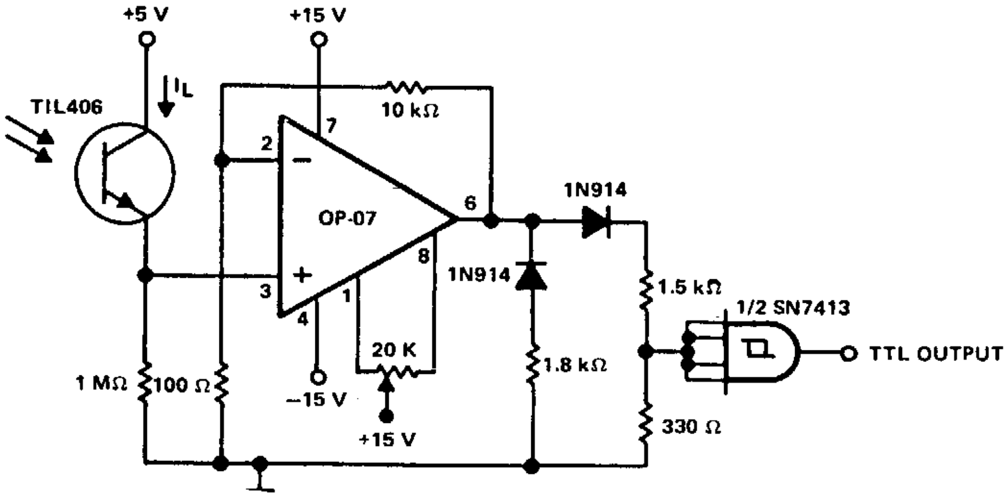
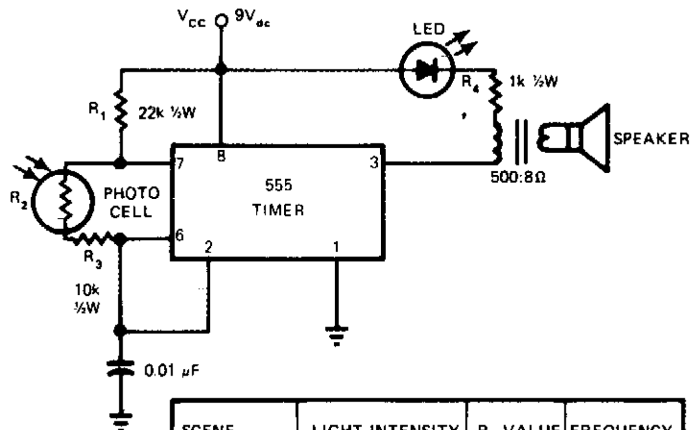


Fig. 52-4

OPTICAL SENSOR-TO-TTL INTERFACE (*Cont.*)

This circuit is designed to detect a low light level at the sensor, amplify the signal, and provide a TTL-level output. When the optical sensor detects low-level light, on condition, its output is small and must be amplified. An amp with very low input bias current and high input resistance must be used to detect the on condition. When sensor TIL406 is in the on condition, its output is assumed to be 250 nA (allowing a safety margin). This results in a 250-mV signal being applied to the noninverting input of amplifier OP-07. Because of the circuit configuration, the OP-07 provides a gain of 100 and its output is in positive saturation. The OP-07 output level is applied to a loading network that provides the basic TTL level.

LIGHT-SENSITIVE AUDIO OSCILLATOR



SCENE	LIGHT INTENSITY	R_2 VALUE	FREQUENCY
BRIGHT SUN	10^4 CANDELA/m^2	100Ω	6500 Hz
DUSK	10^0	10k	3000 Hz
STARLIGHT	10^{-4}	10M	2 - 4 Hz

EDN

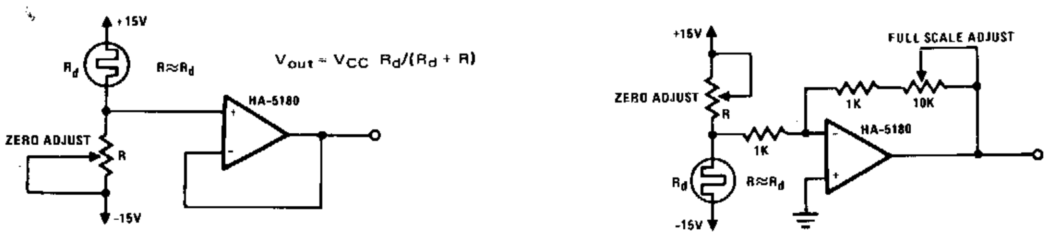
Fig. 52-5

This circuit's frequency of oscillation increases directly with light intensity. The greater the light intensity, the higher the frequency of the oscillator. The 555 timer operates in the astable oscillator mode where frequency and duty cycle are controlled by two resistors and one capacitor. The capacitor charges through R1 and R2, and discharges through R2, a standard photo cell. Resistor R3 limits the upper frequency of oscillation to the audio range. The lower range of approximately 1 pps is set by the value of R2, approximately 10-MΩ, with the photo cell almost totally dark.

A loudspeaker provides audio output, and an LED is used as a pilot light that flashes when the frequency falls below about 10 to 12 Hz. Extremely sensitive, especially on the dark end of the photocell resistance range, the unit can detect lightning many miles away, providing a rapid frequency increase with each flash of lightning. When used with a flashlight at night, the device becomes a simple optical radar for the blind, showing angular direction to a light-reflecting object, as well as height and distance to the object when hand scanned back and forth.

This light-sensitive audio oscillator can also serve as an audible horizontal level device by noting the position of a liquid bubble illuminated by a light source. Thus, you can sense fluid levels as well as the vibration state of a fluid surface level.

LIGHT LEVEL DETECTOR



Cadmium Sulfide cells control two light detection circuits.
(A)

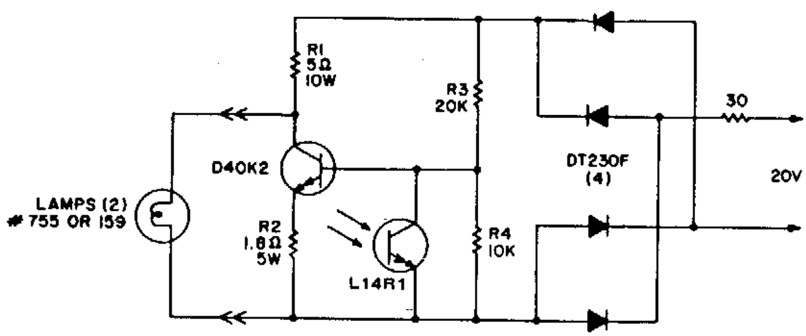
Cadmium Sulfide cells control two light detection circuits.
(B)

HARRIS

Fig. 52-6

If R, the sensor matching resistor, is equal to the "dark" resistance of the cadmium sulfide cell, the amplifier output will range from 0 to $\approx$ 2 as the light level ranges from "dark" to "bright." The circuit in Fig. B operates similarly, but use the standard noninverting configuration instead of the voltage-follower configuration; this allows for variable gain. Although the "dark" resistance of the cadmium sulfide cell is only $\approx$ 7 KΩ, the principles of operation apply to other types of detectors which require the high-input impedance of the HA-5180 for reasonable linearity and useability.

LIGHTED DISPLAY BRIGHTNESS CONTROL



GE

Fig. 52-7

This circuit provides a very low cost method of controlling light levels. Circuit power is obtained from a relatively high source impedance transformer or motor windings, normally used to drive the low-voltage lamps used in these functions. It should be noted that the bias resistors are optimized for the 20-V, 30-Ω source, and they must be recalculated for other sources. The L14R1 is placed to receive the same ambient illumination as the display and should be shielded from the light of the display lamps. The illumination level of lighted displays should be lowered as the room ambient light dims, to avoid undesirable or unpleasant visual effects.

WARNING LIGHT AND MARKER LIGHT

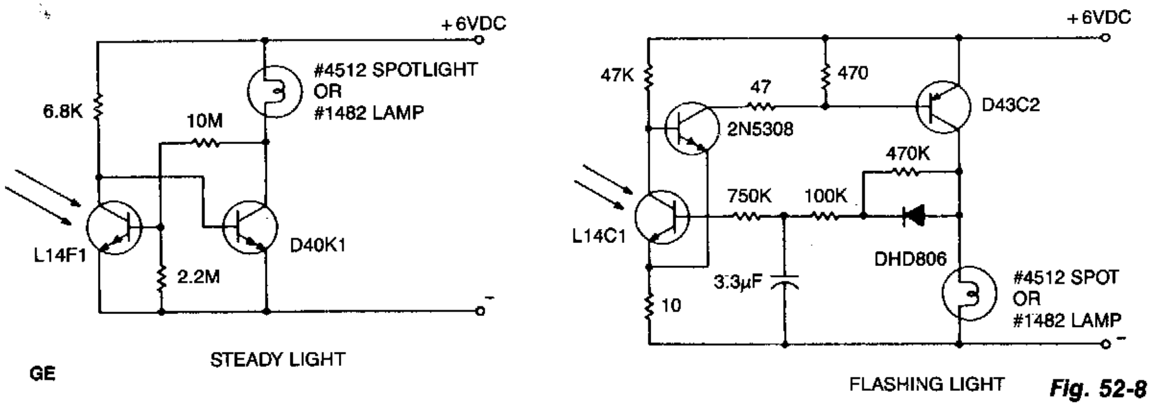


Fig. 52-8

A flashing light of high brightness and short duty cycle is often desired to provide maximum visibility and battery life. This necessitates using an output transistor, which can supply the cold filament surge current of the lamp while maintaining a low saturation voltage. The oscillation period and flash duration are determined in the feedback loop, while the use of a phototransistor sensor minimizes sensitivity variations.

LIGHT-CONTROLLED ONE-SHOT TIMER

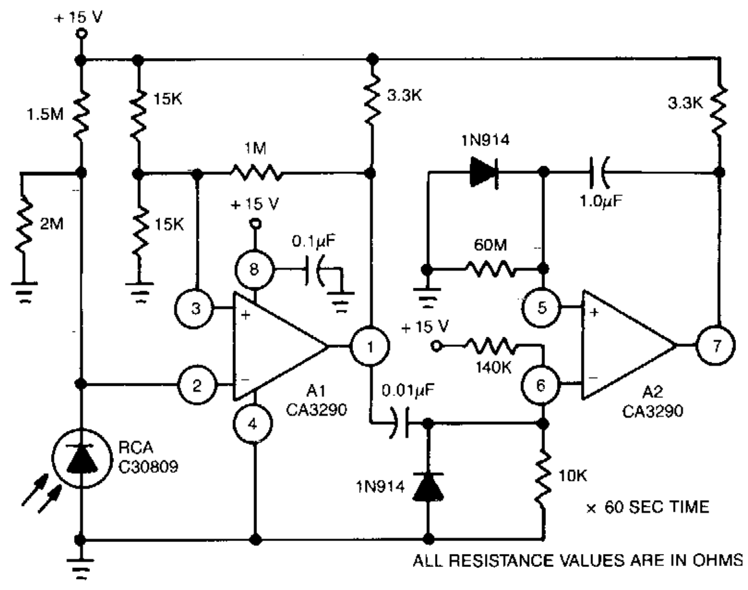
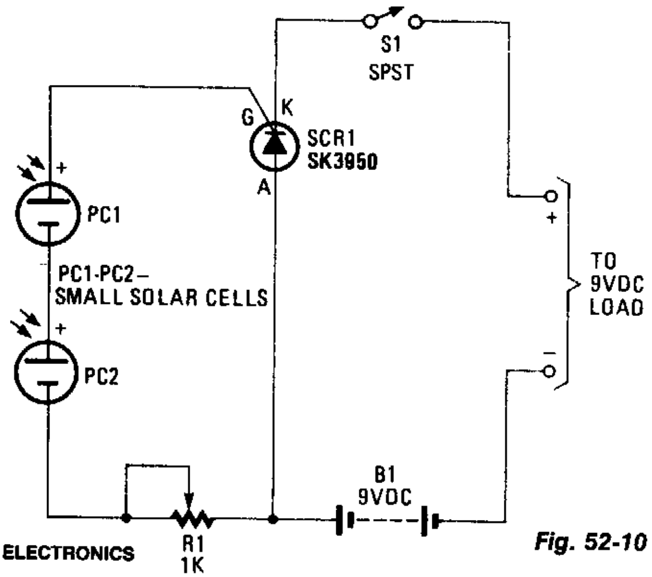


Fig. 52-9

This circuit uses A1 of the CA3290 BiMOS dual voltage comparator to sense a change in light diode current. A2, a one-shot timer, is triggered by the A1 output. If the light source to the photodiode is interrupted, the circuit output switches to a low state for approximately 60 s.

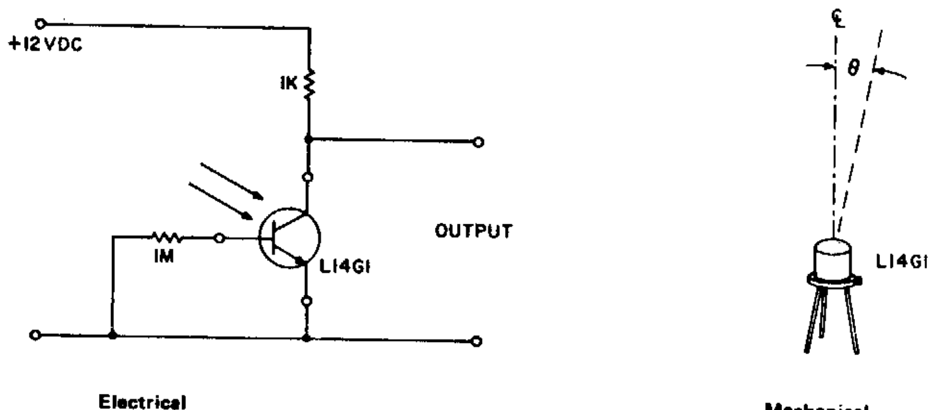
SOLAR-TRIGGERED SWITCH



HANDS-ON ELECTRONICS

Fig. 52-10

SUN TRACKER

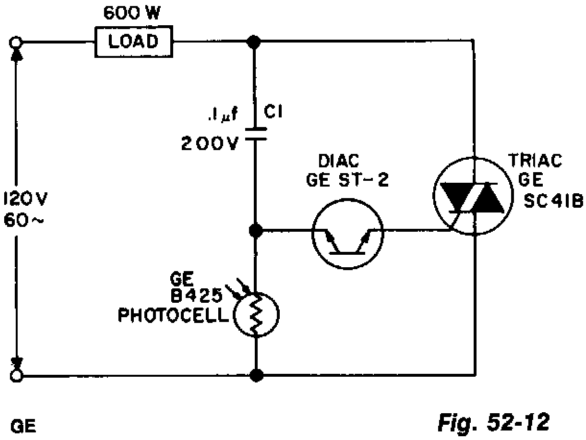


GE

Fig. 52-11

In solar cell array applications and solar instrumentation, it is desirable to monitor the approximate position of the sun to allow efficient automatic alignment. The L14G1 lens can provide about 15° of accuracy in a simple level sensing circuit, and a full hemisphere can be monitored with about 150 phototransistors. The sun provides ≈ 80 mW/cm² to the L14G1 when on the centerline. This will keep the output down to ≤ 0.5 V for $\theta \leq 7.5°$. The sky provides ≈ 0.5 mW/cm² to the L14G1 and will keep the output greater than 10 V when viewed. White clouds viewed from above can lower this voltage to ≈ 5 V on some devices. This circuit can directly drive TTL logic by using the 5-V supply and changing the load resistor to 430 Ω. Different bright objects can also be located with the same type of circuitry simply by adjusting the resistor values to provide the desired sensitivity.

PHOTOELECTRIC AC POWER SWITCH



For a dark photocell, high resistance, the voltage across the diac rises rapidly with the line voltage due to the current through C1, triggering the diac early in the cycle. When the photocell resistance is less than about 2000 Ω, the drop across it is limited to less than the diac triggering voltage, and the load power is shut off.

Fig. 52-12

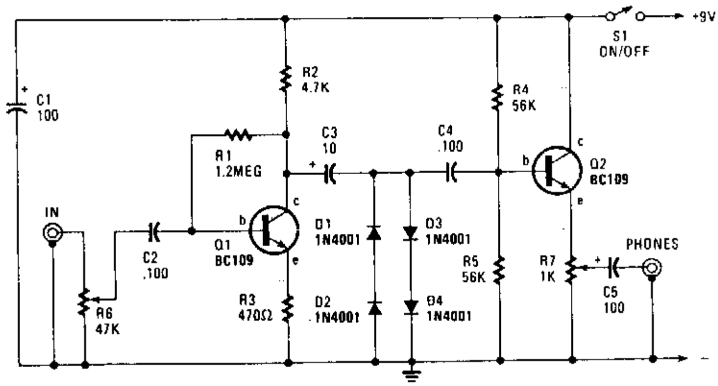
53

Limiters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Noise Limiter
Dynamic Noise Reduction Circuit
Output Limiter

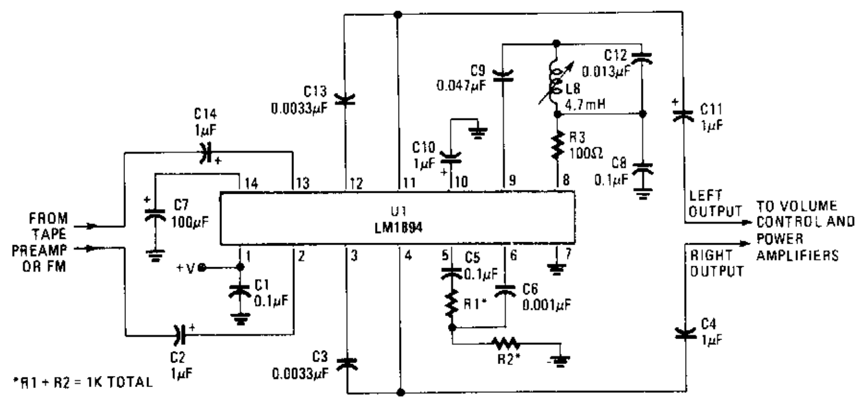
NOISE LIMITER



HANDS-ON ELECTRONICS *Fig. 53-1*

This circuit is fed from the earphone jack of your receiver and goes to limiter control R6 and is then amplified by Q1: a common-emitter stage that has a voltage gain of only about 10, because of the negative feedback introduced by R3. The output of Q1 is fed to a simple clipping circuit, consisting of diodes D1 through D4. The diodes, connected in pairs, act like Zeners with an avalanche rating of about 1 V. The two pairs are connected opposite in polarity to each other, so that the audio signal is clipped at about 1 V. The signal is then coupled to the output socket through an emitter-follower buffer stage built around Q2 and an output attenuator control R7.

DYNAMIC NOISE REDUCTION CIRCUIT



POPULAR ELECTRONICS *Fig. 53-2*

OUTPUT LIMITER

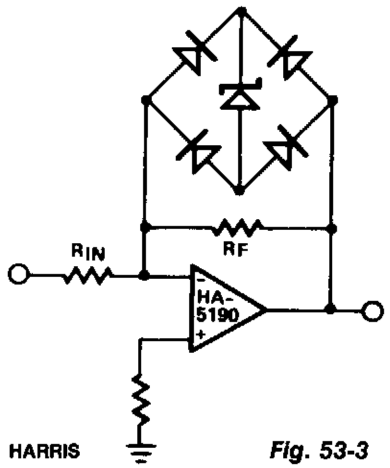


Fig. 53-3

HA-5190 is rated for ±5 V output swing, and saturates at ±7 V. As with most op amps, recovery from output saturation is slow compared to the amplifier's normal response time. Some form of limiting, either of the input signal or in the feedback path, is desirable if saturation might occur. The circuit illustrates a feedback limiter, where gain is reduced if the output exceeds $\pm\,(V_Z + 2V_f)$. A 5-V zener with a sharp knee characteristic is recommended.

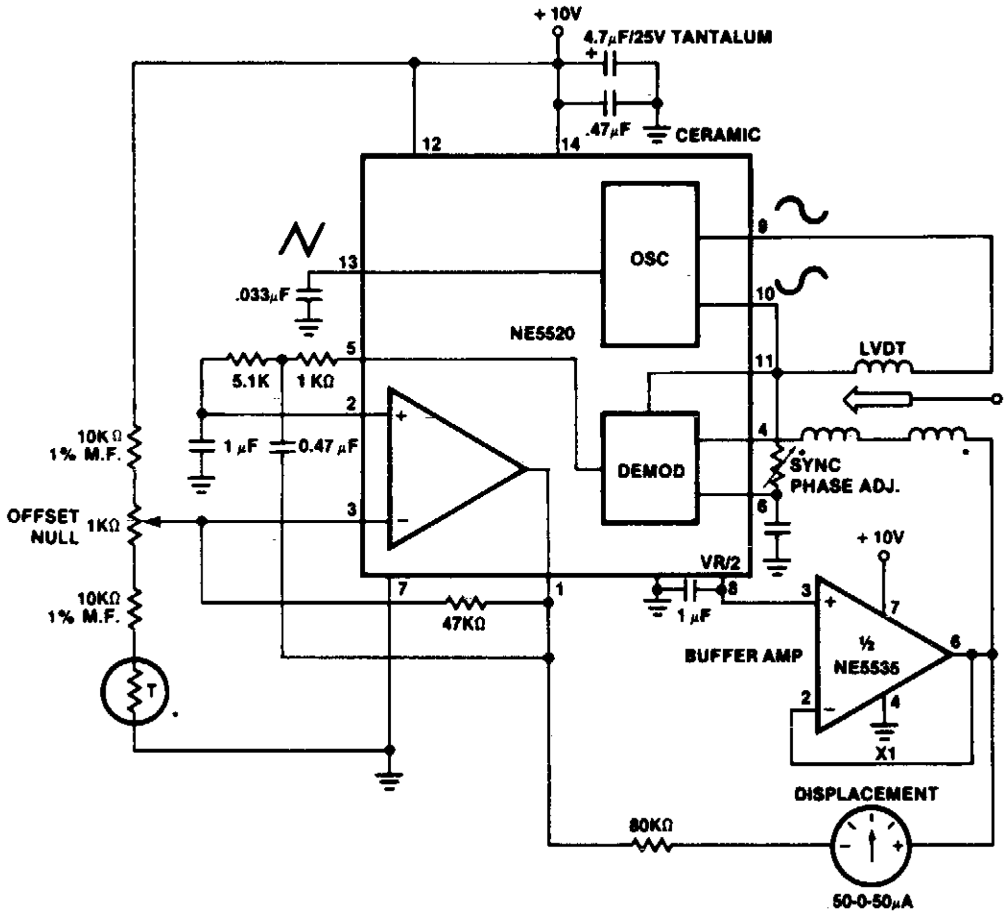
54

LVDT Circuit

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

LVDT Driver Demodulator

LVDT DRIVER DEMODULATOR



SIGNETICS

Fig. 54-1

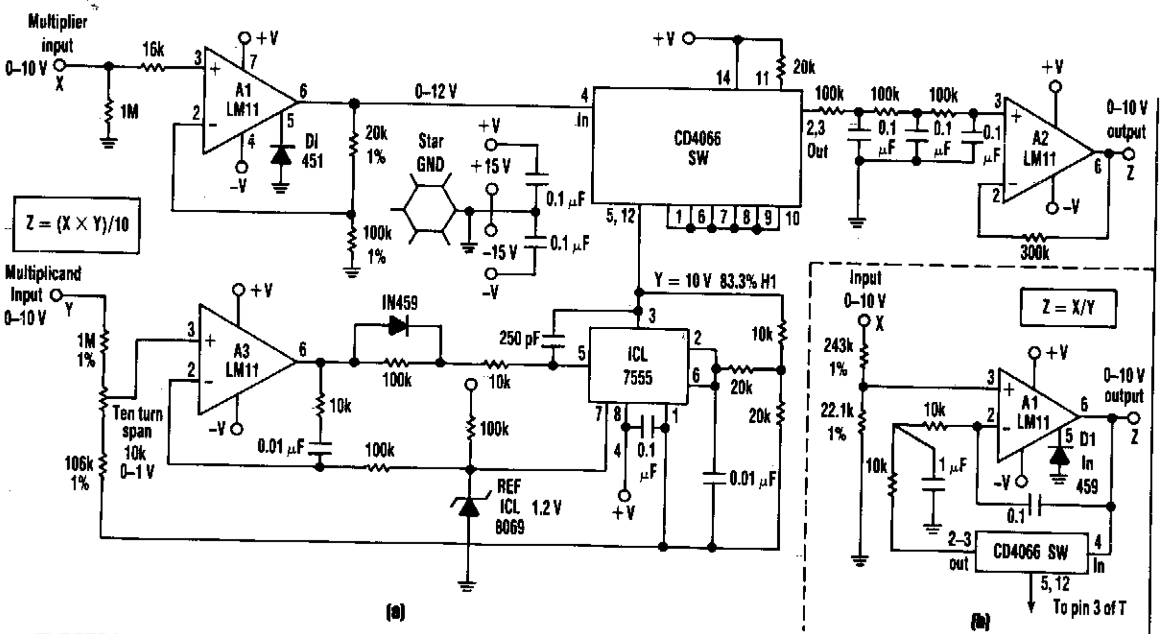
A very simple motion transducer can be constructed using the circuit shown. The output is biased to one-half the supply voltage. This requires special interface circuitry for the signal readout. One simple method is to use a zero center meter in a bridge configuration. Displacement now can be measured as a positive or negative meter reading. Readout sensitivity is a function of the particular LVDT and of the gain of the error amplifier. Dc offsets can be nulled by using a simple offset adjustment circuit as indicated.

55

Mathematical Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DIVIDE/MULTIPLY WITH ONLY ONE TRIM



ELECTRONIC DESIGN

Fig. 55-1

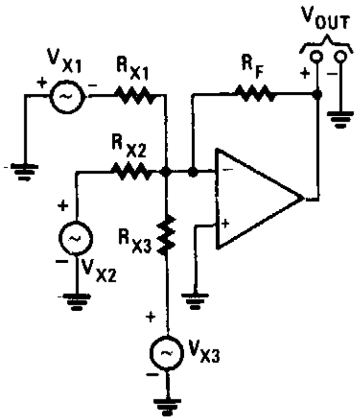
This relatively simple, inexpensive circuit requiring one trimming operation can multiply or divide with a consistent accuracy of greater than 1 part in 1,000. An inexpensive CMOS version of standard 555 timer chip T, in conjunction with low-drift LM11 error amplifier A3, an inexpensive analog chopper switch SW, form a unique voltage-to-duty-cycle converter to produce the difficult transfer function necessary for accurate conversion.

An unknown multiplicand voltage applied to the A3 error op amp circuit's Y input controls the duty cycle of the timer through its pin 5 modulation input. The network between the sink-and-source output of the timer, pin 3, and the state trigger inputs, pins 2 and 6, cause the timer to oscillate. An error feedback signal from the timer's discharge output, pin 7, represents the duty cycle. Integrating this duty-cycle signal with voltage reference REF representing full scale, and applying the result to the inverting input of A3, closes the feedback loop and insures high accuracy.

Multiplier X feeds into another LM11 op amp, A1, which acts as a input buffer and scaler. A third LM11, A2, filters and buffers the Z output. Between A1 and A2, the timer's duty-cycle output modulates the analog switches of a CD4066 to achieve the desired multiplier output. To perform division instead of multiplication, reconfigure the op amp A1 circuit with the use of jumpers. Amplifier A2 isn't required in the division configuration.

To calibrate the circuit, connect the X and Y inputs together and apply 10 V. Then adjust the 10-turn span potentiometer to achieve a 10-V output at Z for multiplication, or 1 V for the division configuration. Also check for zero output at a zero multiplier input. The circuit is scaled for 0 – 10 V inputs and outputs with a small overrage capability, but other scalings are possible. Star grounding or a heavy ground bus should be used to reduce offset problems that are unavoidable in this design.

ADDER

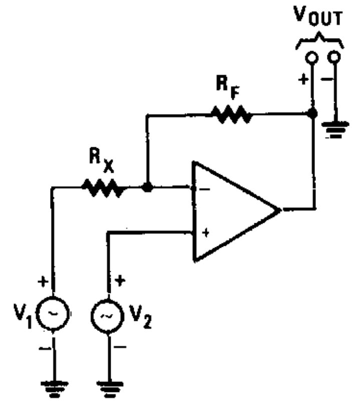


$$V_{OUT} = -\frac{R_F}{R_X}(V_{X1} + V_{X2} + V_{X3})$$
$$(R_{X1} = R_{X2} = R_{X3})$$

HANDS-ON ELECTRONICS

Fig. 55-2

SUBTRACTOR



$$V_{OUT} = V_2 + \frac{R_F}{R_X}(V_2 - V_1)$$

HANDS-ON ELECTRONICS

Fig. 55-3

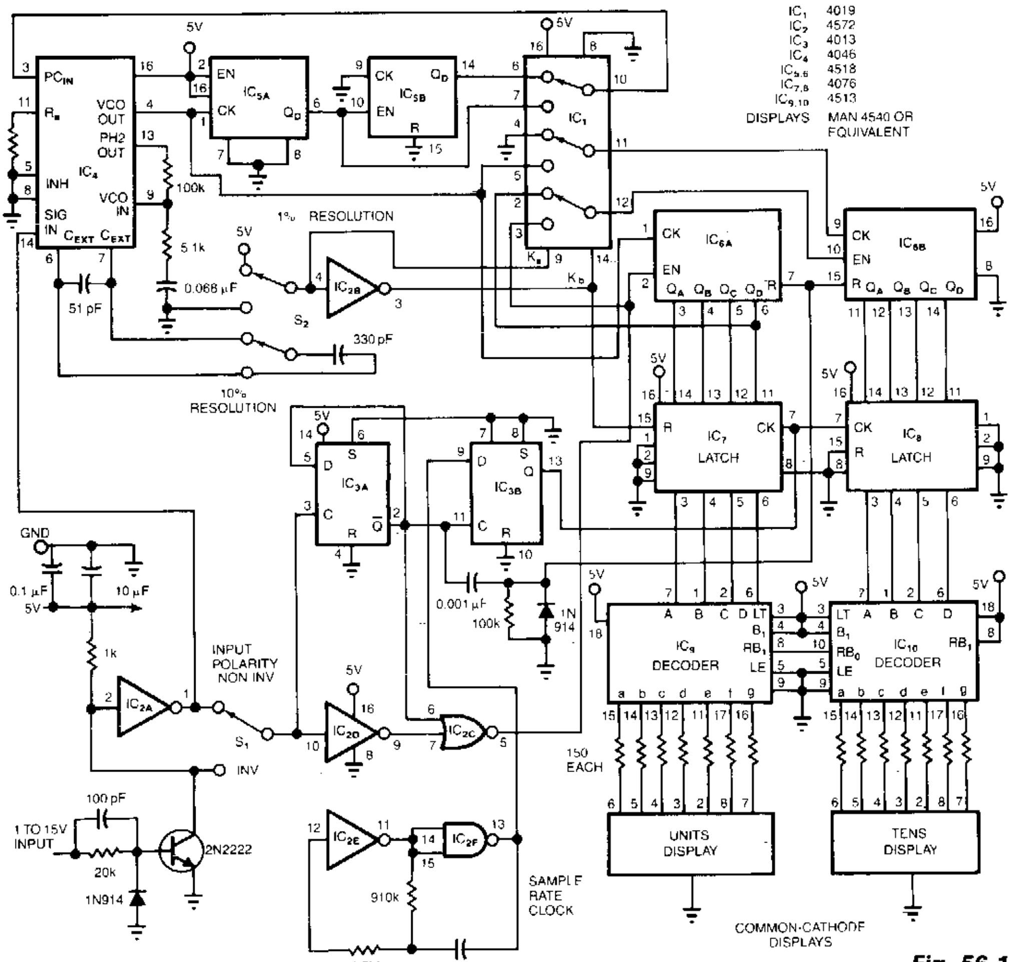
56

Measuring and Test Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

- Duty Cycle Monitor
- 3-in-1 Test Set
- Stereo Power Meter
- Wide-Range Rf Power Meter
- LED Peak Meter
- Lc Checker
- Tachometer and Direction-of-Rotation Circuit
- Very Short Pulse-Width Measurer
- QRP SWR Bridge
- Electrostatic Detector
- Current Monitor and Alarm
- Picoammeter Circuit
- Paper Sheet Discriminator for Printing and Copying Machines
- Precision Frequency Counter/Tachometer
- Motor Hour Meter
- Stud Finder
- Low-Power Magnetic Current Sensor
- Line-Current Monitor
- S Meter
- Hot-Wire Anemometer
- Audible Logic Tester
- SCR Tester
- Digital Frequency Meter
- Low-Current Measurement System
- Simple Continuity Tester
- Sound-Level Meter
- LED Panel Meter
- Optical Pick-Up Tachometer
- Peak-dB Meter

DUTY CYCLE MONITOR

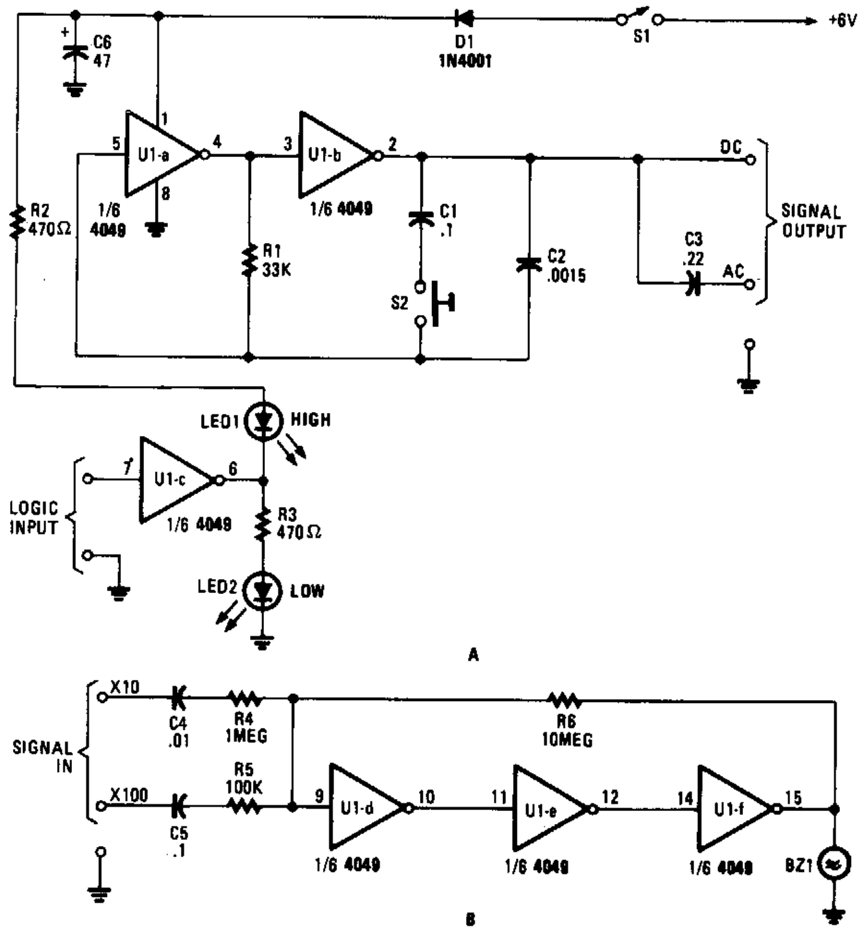


EDN

Fig. 56-1

The circuit monitors and displays a digital signal's duty cycle and provides accuracy as high as ±1%. Using switch S2, you can choose a frequency range of either 250 Hz to 2.5 kHz at ±1% accuracy or 2 kHz to 50 kHz at ±10% accuracy. The common-cathode display gives the signal's duty-cycle percentage. Phase-locked loop IC4 and counters IC5A and IC5B multiply the input frequency by a factor of either 10 or 100, depending on switch S2's setting. IC6A and IC6B count this multiplied frequency during the incoming signal's mark interval. IC7 and IC8 then latch this count and display it at the clock's sample rate. For example, if you select a 1% resolution, when the signal's mark period is 40% of the total period, the circuit will enable the counter comprising IC6A and IC6B for 40 counts. To obtain space-interval sampling, you can reverse the input polarity using switch S1. IC2A samples the input signal's period and enables gate IC2C and resets the counter. IC2E and IC2F form the sample-rate clock; IC3B synchronizes the clock's output with the input, so that the circuit can update latches IC7 and IC8.

3-IN-1 TEST SET



POPULAR ELECTRONICS

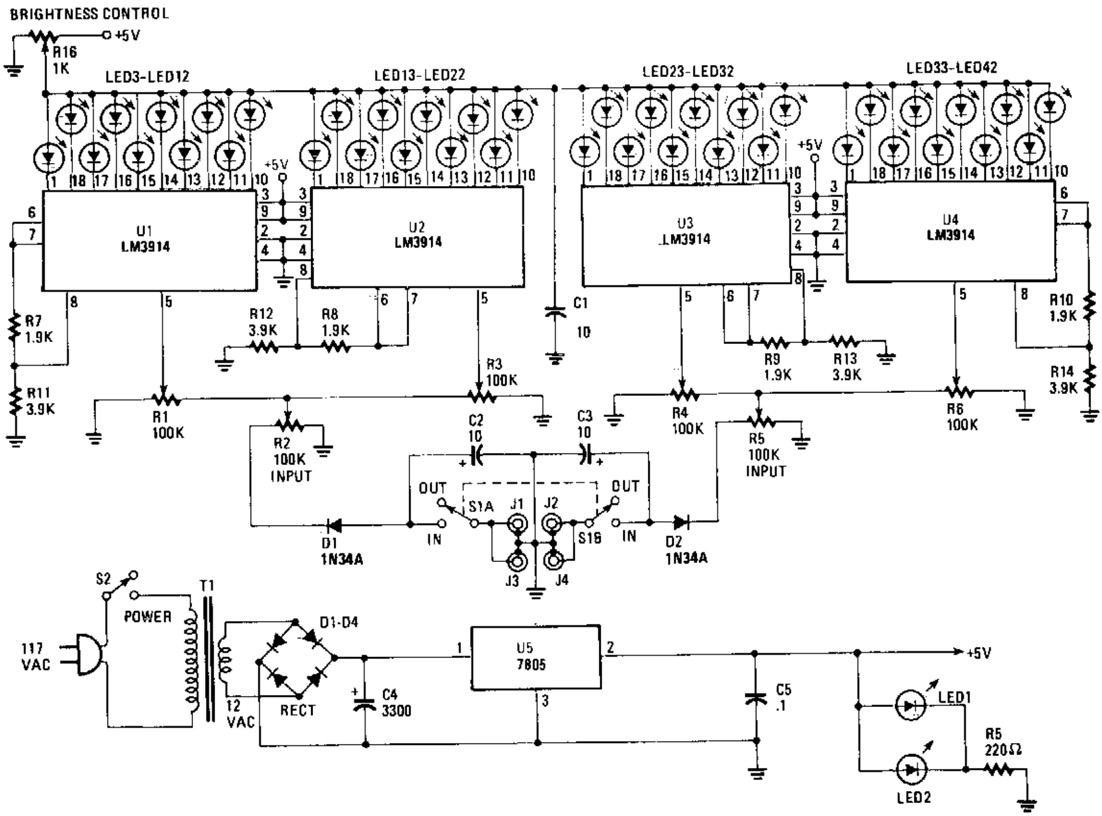
Fig. 56-2

This circuit is designed around a 4049 hex inverter/buffer. Two inverters are used in a dual-frequency signal-injector circuit, another inverter is used as a logic probe, and the remaining three inverters are used as a sensitive dual-input, audio-signal tracer.

The signal-injector portion gates are configured as a two-frequency, pulse-generator circuit. Under normal conditions, the generator's output frequency is around 10 kHz, but when S2 is closed, the output frequency drops to about 100 Hz. The logic-probe portion is made up of U1c, the output of the inverter decreases. The low output of U1c reverse biases LED2, so it remains off. That low output also forward biases LED1, causing it to light. But when a logic low presented U1c's input, the situation is reversed, so LED2 lights and LED1 darkens.

The audio-signal tracer portion is made up of the three remaining inverters which are configured as a linear audio amplifier to increase the input signal level by a factor of 10 or 100. The amplified output signal feeds a miniature piezo element of audible detection.

STEREO POWER METER

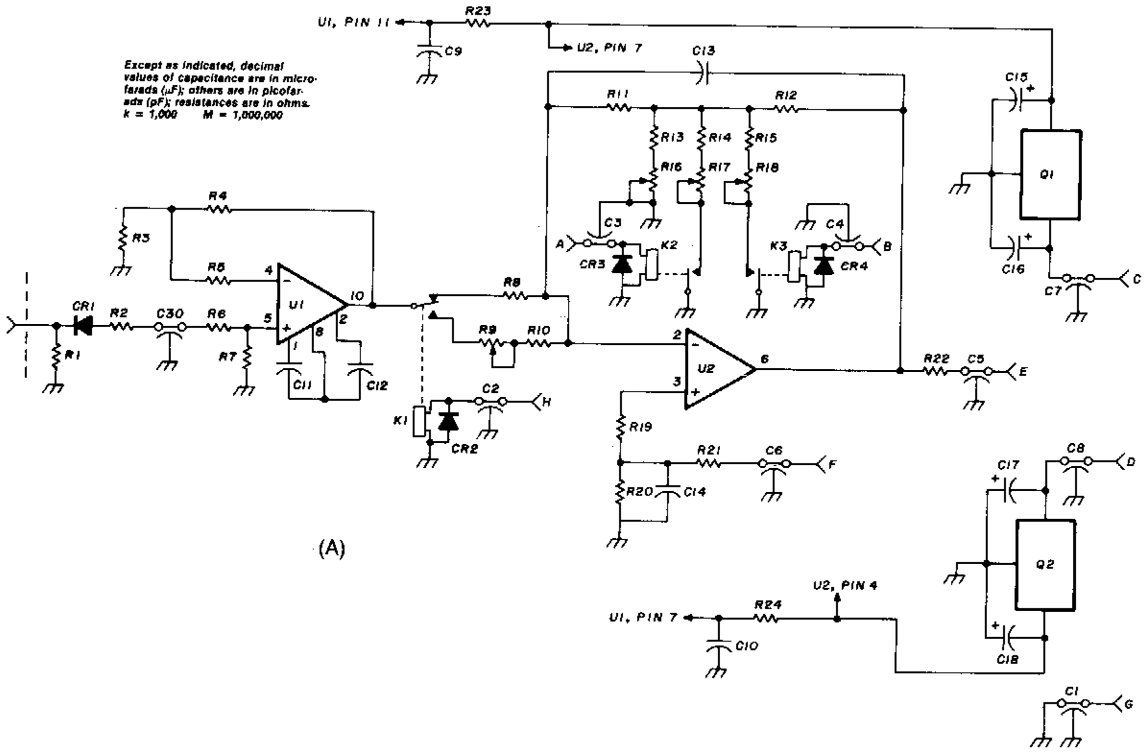


HANDS-ON ELECTRONICS

Fig. 56-3

The Stereo Power Meter is made up of two identical circuits and a power supply. Each circuit contains two LM3914 display chips which contain 10 voltage comparators, a 10-step voltage divider, a reference-voltage source, and a mode-select circuit that selects a bar or dot display via a logic input at pin 9. The brightness of the LEDs is controlled by the 1900-Ω resistors and the reference voltage is controlled by the 3900-Ω resistors. The 10-step voltage divider within the chips is connected between the reference voltage and ground. Since each step of the voltage divider is separated by a 1-KΩ resistor, each comparator senses a voltage 10% greater than the preceding comparator. The signal is applied to pin 5, which is buffered through a resistor-diode network and then amplified as it passes to each of the 10 comparators. Each LED is grounded through the comparators as the input signal voltage matches the reference voltage. That results in one to 10 LEDs illuminating as the signal voltage increases.

WIDE-RANGE RF POWER METER



(A)

table 1. RF power meter and power supply parts list

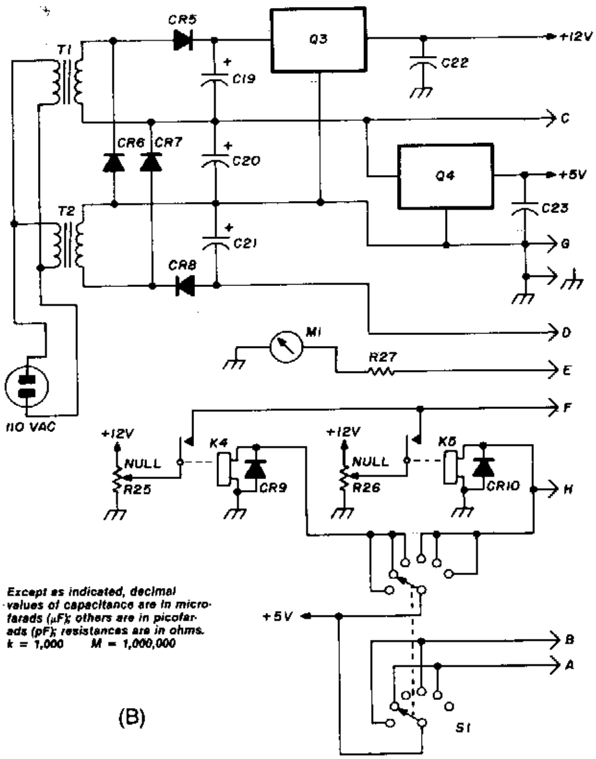
C1 thru C8	1000 pF feedthru (Erie, Cambion)
C9,10,15,18	1 μF 10wvdc tantalum
C11-12	0.1 μF metalized film
C13	500 pF disc
C14	0.01 μF disc ceramic
C16,17	2.2 μF 25 wvdc tantalum
C19,21	100 μF 15 wvdc electrolytic
C20	500 μF 15 wvdc electrolytic
C22,23	0.01 μF disc
C30	100pF chip capacitor
CR1	HSCH-3486 Hewlett-Packard
CR2,3,4,9,10	1N914 or equivalent
CR5,6,7,8	1N4003 or equivalent
K1	SPDT reed Magnecraft W172-DIP5 (internal diode — CR2 not used)
K2,3,4,5	SPST reed EAC EAC Z610-ND
M1	1 mA DC meter with dB scale
Q1,4	78LO5 regulator
Q2	79LO5 regulator
Q3	78L12 regulator
R1,2	50 ohm 1/8 watt carbon film

All resistors 1% metal film 1/4 watt	
R3,6,14,22	1k
R5,7	100k
R10	120k
R4	150k
R8,19	4.99k
R11,12	20k
R13	2.74k
R15	165 ohm
All resistors 5% carbon film 1/4 watt	
R20	100 ohm
R21	1 megohm
R23, 24	10 ohm
R27	1.5k
R9	50k Panasonic CEG54 trimpot
R16	500 ohm Panasonic CEG52 trimpot
R17	200 ohm Panasonic CEG22 trimpot
R18	100 ohm Panasonic CEG12 trimpot
R25,26	10k potentiometer
S1	DP6T rotary switch
T1,T2	6.3 VAC transformers
U1	ICL76508CPD Intersil
U2	LM11CLH National
Box	CU-124 BUD
Chassis	9 1/2 x 5 x 2 chassis BUD Ac-403

HAM RADIO

Fig. 56-4

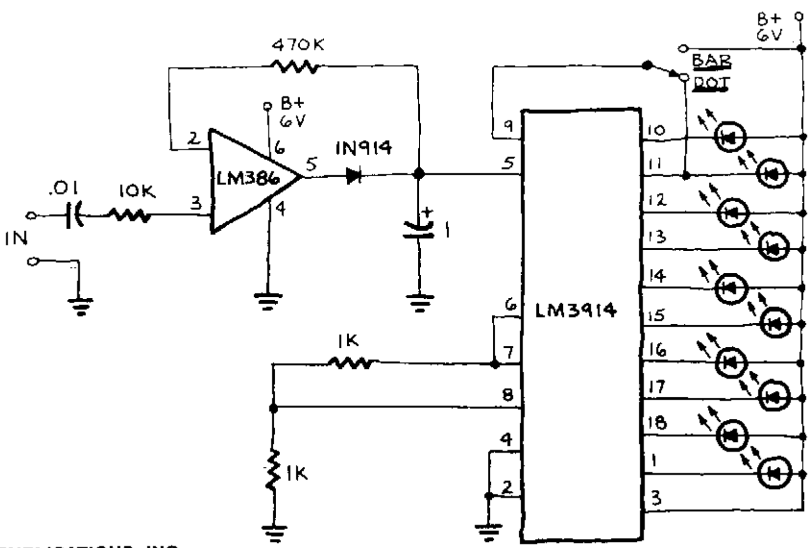
WIDE-RANGE RF POWER METER (*Cont.*)



(B)

The Hewlett-Packard HSCH-3486 zero-bias Schottky diode is used as the detector. To avoid using a modulation method of detection, a chopper-stabilized op amp is used. The chopper op amp basically converts the input dc voltage to ac, amplifies it, and converts it back to dc. Amplifying the dc output from the detector 150 times with a chopper op amp puts the signal at a level that simpler op amps, such as the LM11, can handle. Offset voltages in the amplifier are nulled with two pots, one for the high range and one for the lower three ranges.

LED PEAKMETER

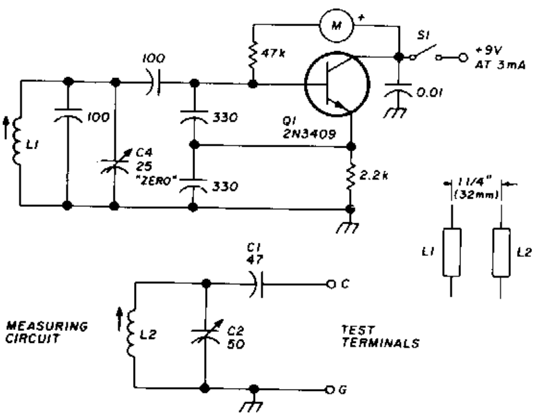


GERNSBACK PUBLICATIONS, INC.

Fig. 56-5

The circuit includes a peak detector that immediately drives the readout to any new higher signal level and slowly lowers it after the signal drops to zero. The readout is a moving dot or expanding bar display. The circuit can be expanded for a longer bar readout. Tapping five or more LED peakmeters into a frequency equalizer or series of audio filters should give a unique result. The bottom LED remains on with no signal at the input, thus providing a pilot light for the unit.

LC CHECKER



L1 - 30T. NO. 28E CLOSE-WOUND ON 3/8" (10mm) SLUG-TUNED FORM. APPROX. 7μH.
L2 - 50T. SAME AS L1. APPROX. 30μH.
Q1 - 2N3904 OR SIMILAR.
M - 0 TO 100 OR 0 TO 200μA.
S1 - SPST TOGGLE OR SLIDE SWITCH.

HAM RADIO ***Fig. 56-6***

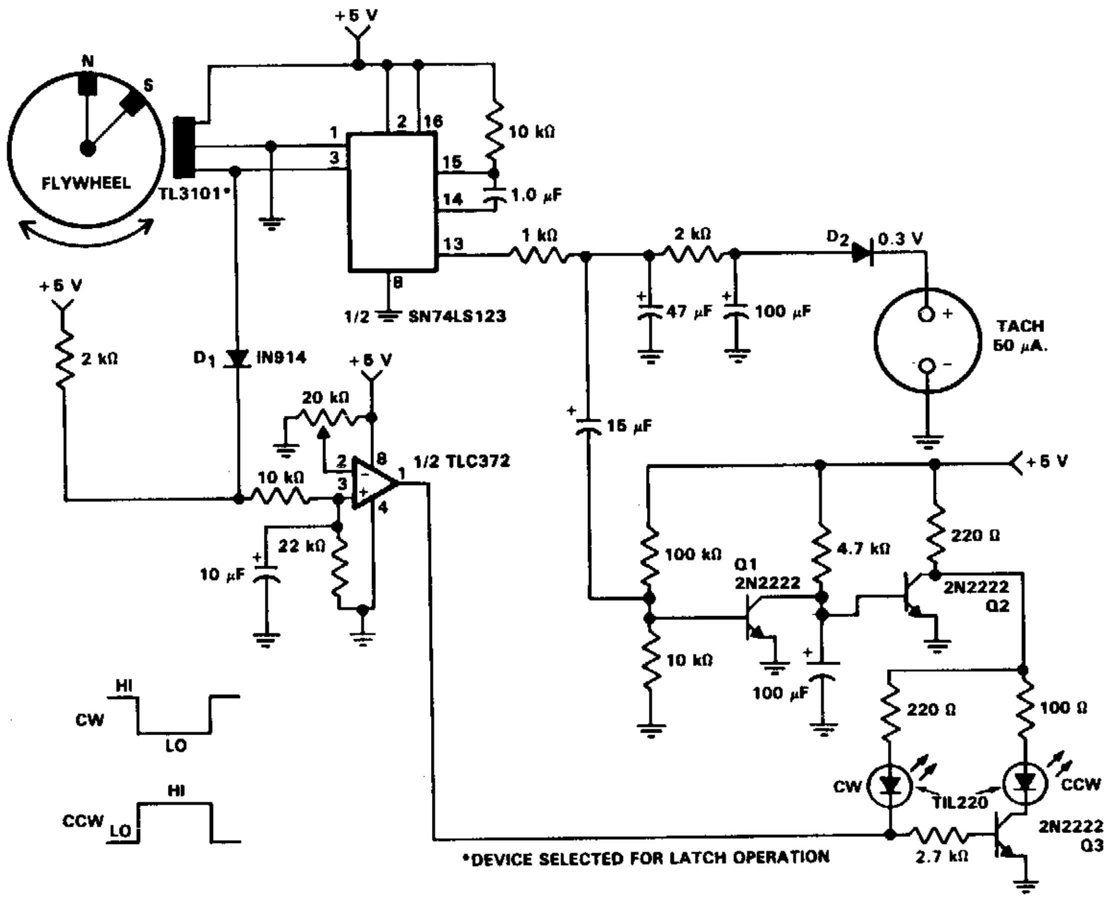
The circuit is based on the *grid-dip* or *absorption effect,* which occurs when a parallel resonant circuit is coupled to an oscillator of the same frequency. Q1 operates in a conventional Colpitts oscillator circuit at a fixed frequency of approximately 4 MHz. A meter connected in series with the transistor's base-bias resistor serves as the dip or absorption indicator.

The variable measuring circuit consists of C1, C2, and L2 and is connected to panel terminals as shown. L2 is loosely coupled to L1 in the oscillator circuit. This measuring circuit is tuned to the oscillator frequency with variable capacitor C2 set at full capacitance. When power is applied to the oscillator, the meter shows a dip caused by power absorption from the measuring circuit.

Connecting an unknown capacitor across the test terminals lowers the resonant frequency of the measuring circuit. To restore resonance, tune capacitor C2 lower in capacitance. The meter will dip again when you reach this point. Determine the capacitance across the text terminals by calibrating the dial settings of C2.

Capacitor C4, a small variable trimmer in the oscillator circuit, compensates for drift or other variations and is normally set at half capacitance. The capacitor is a panel control, labeled zero, and it is used to set the oscillator exactly at the dip point when C2 is set at maximum capacitance. This corresponds to zero on the calibration scale.

TACHOMETER AND DIRECTION-OF-ROTATION CIRCUIT



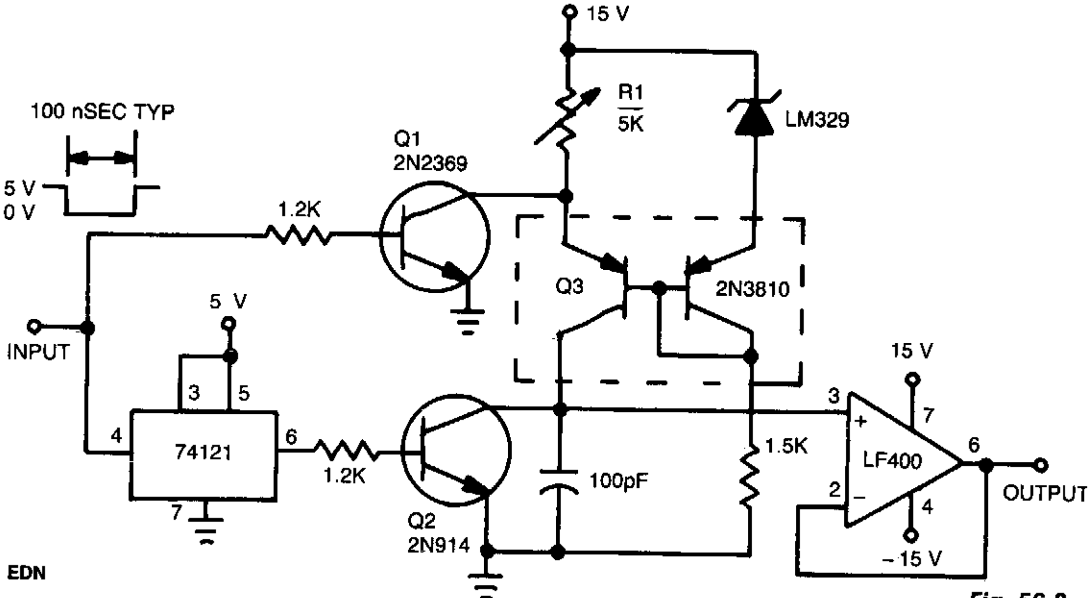
Reprinted by permission of Texas Instruments.

Fig. 56-7

In machine and equipment design, some applications require measurement of both the shaft speed and the direction of rotation. Fig. 56-4 shows the circuit of a tachometer, which also indicates the direction of rotation. The flywheel has two magnets embedded in the outer rim about 45° apart. One magnet has the north pole toward the outside and the other magnet has the south pole toward the outside rim of the flywheel. Because of the magnet spacing, a short on pulse is produced by the TL3101 in one direction and a long on pulse in the other direction. A 0 – 50 μA meter is used to monitor the flywheel speed while the LEDs indicate the direction of rotation. The direction-of-rotation circuit can be divided into three parts:

- TLC372 device for input conditioning and reference adjustment.
- Two 2N2222 transistors which apply the V_{CC} to the two LEDs when needed.
- The two TIL220 LEDs which indicate clockwise (CW) or counterclockwise (CCW) direction of rotation.

VERY SHORT PULSE-WIDTH MEASURER

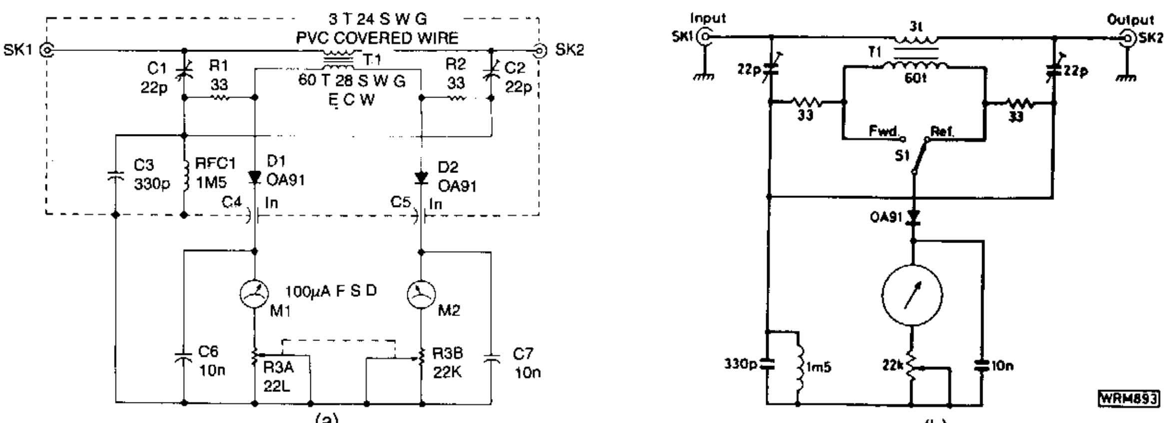


EDN

Fig. 56-8

This circuit operates by charging a small capacitor from a constant-current source when the pulse to be measured is present. Dual pnp transistor Q3 is the current source; its output current equals the LM329 reference voltage divided by the resistance of potentiometer R1. When the input is high with no pulse present, Q1 keeps the current source turned off. When the pulse begins and the input decreases, Q1 turns off and the monostable multivibrator generates a short pulse. The pulse from the multivibrator turns on Q2, removing any residual charge from the 100-pF capacitor. Q2 then turns off, and the capacitor begins to charge linearly from the current source. When the input pulse ends, the current source turns off, and the voltage on the capacitor is proportional to the pulse width.

QRP SWR BRIDGE

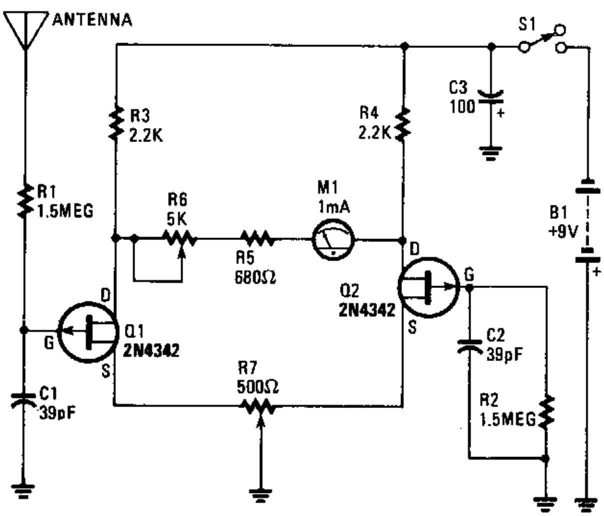


73 AMATEUR RADIO

Fig. 56-9

The design shown is a simple unit for QRP operation on all authorized frequencies up to 30 MHz, based on a toroidal transformer T1. The secondary winding of T1 samples a small amount of rf power, both forward and reflected, which is divided by the bridge circuit and rectified by diodes D1 and D2. Forward and reflected readings are obtained simultaneously on the two meters M1 and M2, and the bridge is matched and balanced at the required load impedance by C1 and C2. See Fig. 56-9b for an alternative, less expensive, single meter version. The bridge also measure forward power.

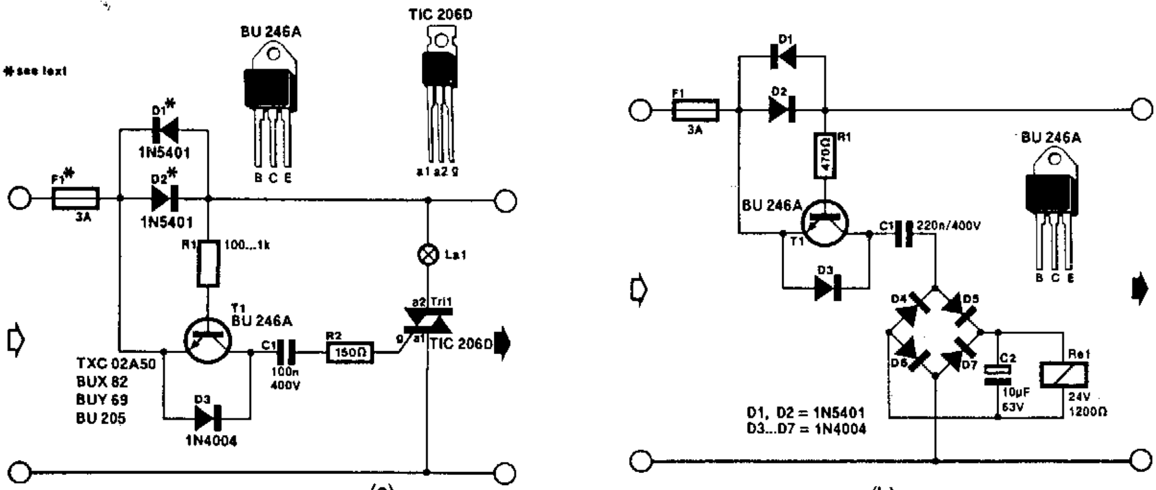
ELECTROSTATIC DETECTOR



HANDS-ON ELECTRONICS *Fig. 56-10*

The heart of the electroscope is the two junction FETs Q1 and Q2 connected in a balanced-bridge circuit. The gate input of Q1 is connected to the wire pick-up antenna, while Q2's gate is tied to the circuit's common ground through R2. That type of bridge circuit offers excellent temperature stability; therefore, Q1 is allowed to operate in an open-gate configuration. Potentiometer R7 is used to balance the bridge circuit, and R6 sets the maximum meter swing. Capacitors C1 and C2 help to reduce the 60-Hz pickup and add to the short-term stability of the circuit.

CURRENT MONITOR AND ALARM



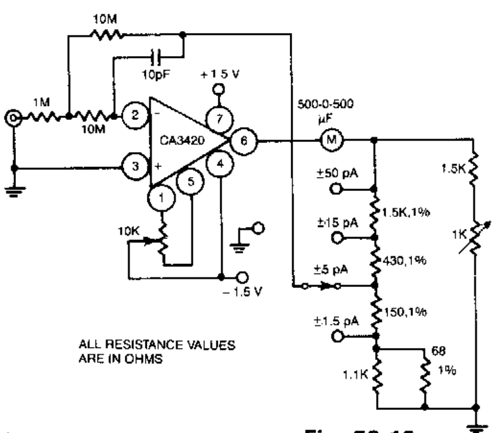
ELEKTOR ELECTRONICS

Fig. 56-11

The circuit in Fig. 56-11a lights the signal lamp upon detecting a line current consumption of more than 5 mA, and handles currents of several amperes with appropriate diodes fitted in the D1 and D2 positions. Transistor T1 is switched on when the drop across D1/D2 exceeds a certain level. Diodes from the well-known 1N400x series can be used for currents of up to 1 A, while 1N540x types are rated for up to 3 A. Fuse F1 should suit the particular application.

The circuit in Fig. 56-11b is a current-triggered alarm. Rectifier bridge D4 through D7 can only provide the coil voltage for Rel when the current through D1/D2 exceeds a certain level, because then series capacitor C1 passes the alternating main current. Capacitor C1 needs to suit the sensitivity of the relay coil. This is readily effected by connecting capacitors in parallel until the coil voltage is high enough for the relay to operate reliably.

PICOAMMETER CIRCUIT

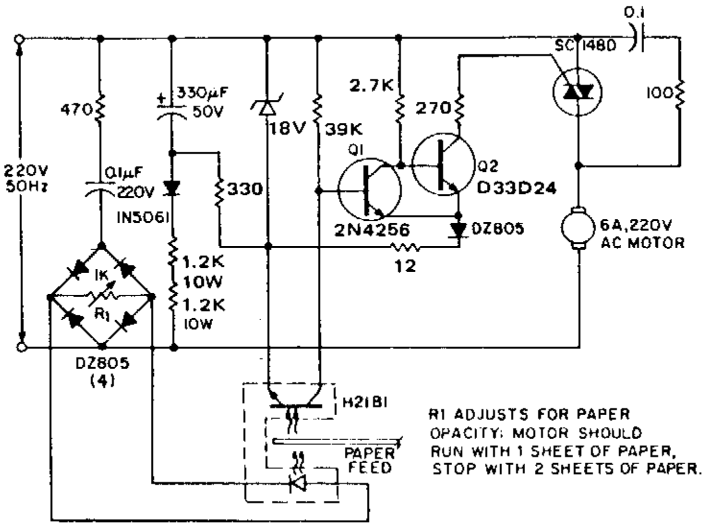


GE/RCA

Fig. 56-12

This circuit uses the exceptionally low input current 0.1 pA of the CA3420 BiMOS op amp. With only a single 10-MΩ resistor, the circuit covers the range from ±50 pA to a maximum full-scale sensitivity of ±1.5 pA.

PAPER SHEET DISCRIMINATOR FOR PRINTING AND COPYING MACHINES

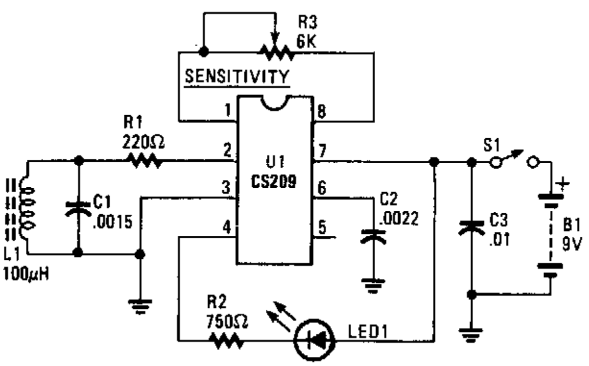


GE

Fig. 56-13

The circuit outputs power to the drive motor when one or no sheets are being fed, but interrupts motor power when two or more superimposed sheets pass through the optodetector slot. The optodetector can be either an H2aB darlington interruptor module or an H23B matched emitter-detector pair. The output from the optodarlington is coupled to a Schmitt trigger, comprising transistors Q1 and Q2 for noise immunity and minor paper opacity variation immunity. When the Schmitt is on, gate current is applied to the SC148D output device. The dc power supply for the detector and Schmitt is a simple rc diode half-wave configuration chosen for its low cost (fewer diodes and no transformers) and minimum bulk. While such a supply is directly coupled to the power triac, this is precluded by current drain considerations (50 mA dc for the gate drive alone). Note that direct coupling of the Schmitt to the output triacs is preferred, since RFI is virtually eliminated with the quasi-dc gate drive.

STUD FINDER



HANDS-ON ELECTRONICS

Fig. 56-14

The CS209 is designed to detect the presence or proximity of magnetic metals. It has an internal oscillator that, along with its external lc resonant circuit, provides oscillations whose amplitude is dependent upon the *Q* of the lc network. Close proximity to magnetic material reduces the *Q* of the tuned circuit, thus the oscillations tend to decrease in amplitude. The decrease in amplitude is detected and used in turn on LED1, indicating the presence of a magnetic material (i.e., nail or screw).

PRECISION FREQUENCY COUNTER/TACHOMETER

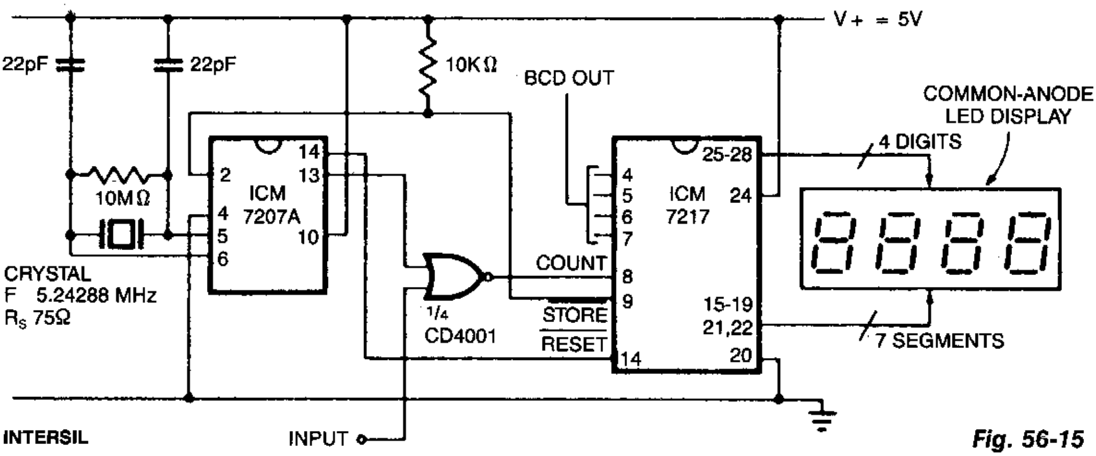


Fig. 56-15

In this configuration, the display reads hertz directly. With pin 11 of the ICM7027A connected to V_{DD}, the gating time will be 0.1 second; this will display tens of hertz as the least significant digit. For shorter gating times, an ICM7207 can be used with a 6.5536-MHz crystal, giving a 0.01 second gating with pin 11 connected to V_{DD}, and a 0.1 second gating with pin 11 open.

MOTOR HOUR METER

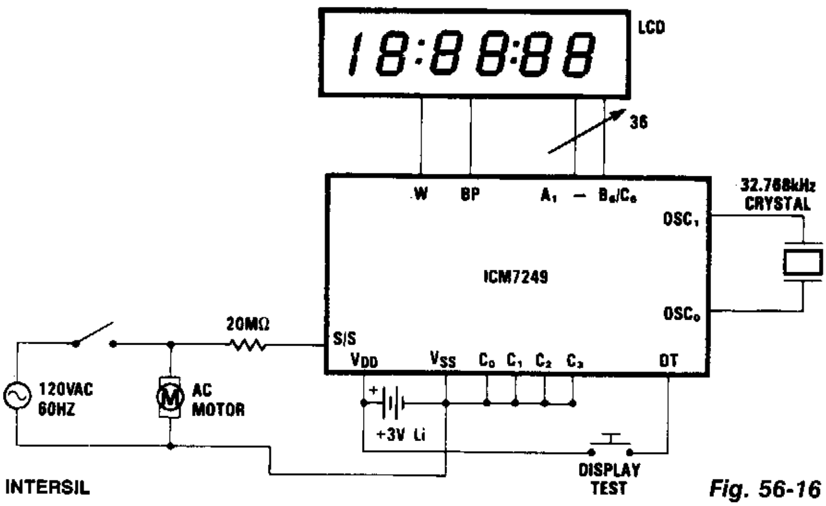
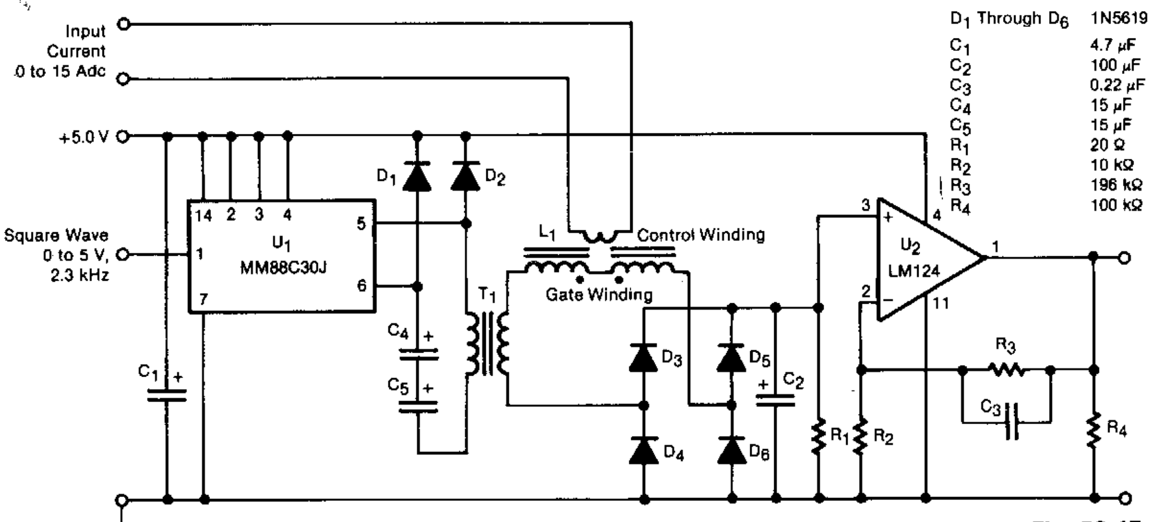


Fig. 56-16

In this application, the ICM7249 is configured as an hours-in-use meter and shows how many whole hours of line voltage have been applied. The 20-MΩ resistor and high-pass filtering allow ac line activation of the S/S input. This configuration, which is powered by a 3-V lithium cell, will operate continuously for 2 1/2 years. Without the display, which only needs to be connected when a reading is required, the span of operation is extended to 10 years.

LOW-POWER MAGNETIC CURRENT SENSOR



NASA

Fig. 56-17

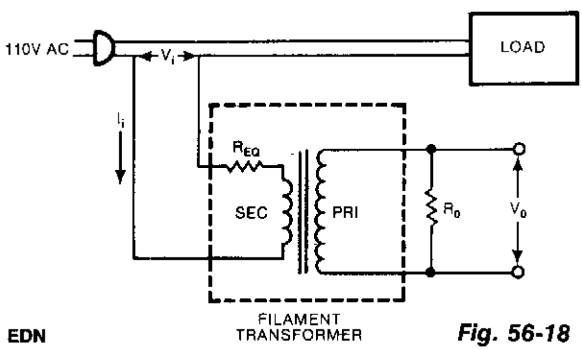
A transducer senses a direct current magnetically, providing isolation between the input and the output. The detecting-and-isolating element is a saturable reactor, in which the input current, to be measured, passes through a one-turn control coil. The transducer provides an output of 0 to 3 Vdc, an input current of 0 to 15 Adc, and consumes 22 mW at 10 Adc input.

Line driver U1 excites the saturable reactor L1 by feeding a 2.3-kHz square wave through transformer T1. The output of L1 is rectified by the bridge rectifier composed of diodes D3 through D6, then amplified by op amp U2, which has a gain of 20.

Diodes D1 and D2 commutate the reactive current fed back to the primary of T1 from L1. Without these diodes, large reactive voltage spikes on the primary would waste power and could destroy U1. Filter capacitor C1 stores the energy fed back through D1 and D2.

To minimize core losses, the core of T1 is made of an alloy of 80% nickel and 20% iron. To minimize capacitance, the primary and secondary windings are interleaved and progressively wound 350°. The primary and secondary windings consist of 408 and 660 turns, respectively, of #34 wire.

LINE-CURRENT MONITOR

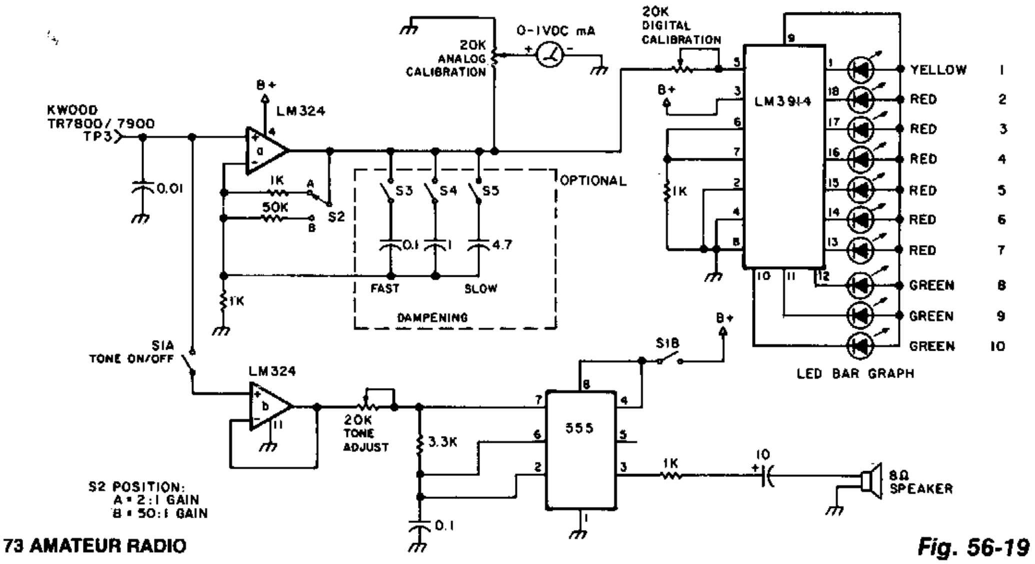


EDN

Fig. 56-18

A low-cost filament transformer provides a linear indication of the load current in an ac line. This method causes a slight series voltage drop over a wide range of load currents.

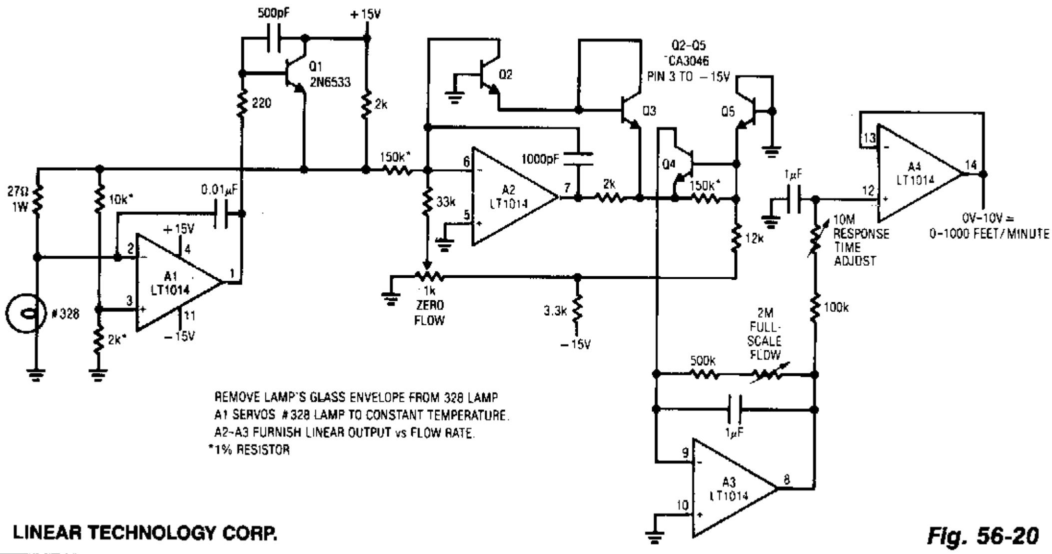
S METER



73 AMATEUR RADIO **Fig. 56-19**

This design is for an external signal strength meter that is analog, digital, and audible for mobile transmitter hunters. The S meter also incorporates a gain circuit. The digital LED bar graph display has a very fast response time. The 3.3-KΩ resistor near LM3914 can be replaced with a 5-K pot to control LED brightness. The S2A position gives a 2:1 gain and the S2B position gives about a 50:1 gain. The calibration pots control the amount of meter action relative to the gain. The optional dampening circuit is used for the averaging of a transmitted signal that has modulated power or when a dip on the voice peaks occur. The capacitors can be switched one by one or switched into a very slow response using 5.8 μF total capacitance.

HOT-WIRE ANEMOMETER



LINEAR TECHNOLOGY CORP. **Fig. 56-20**

AUDIBLE LOGIC TESTER

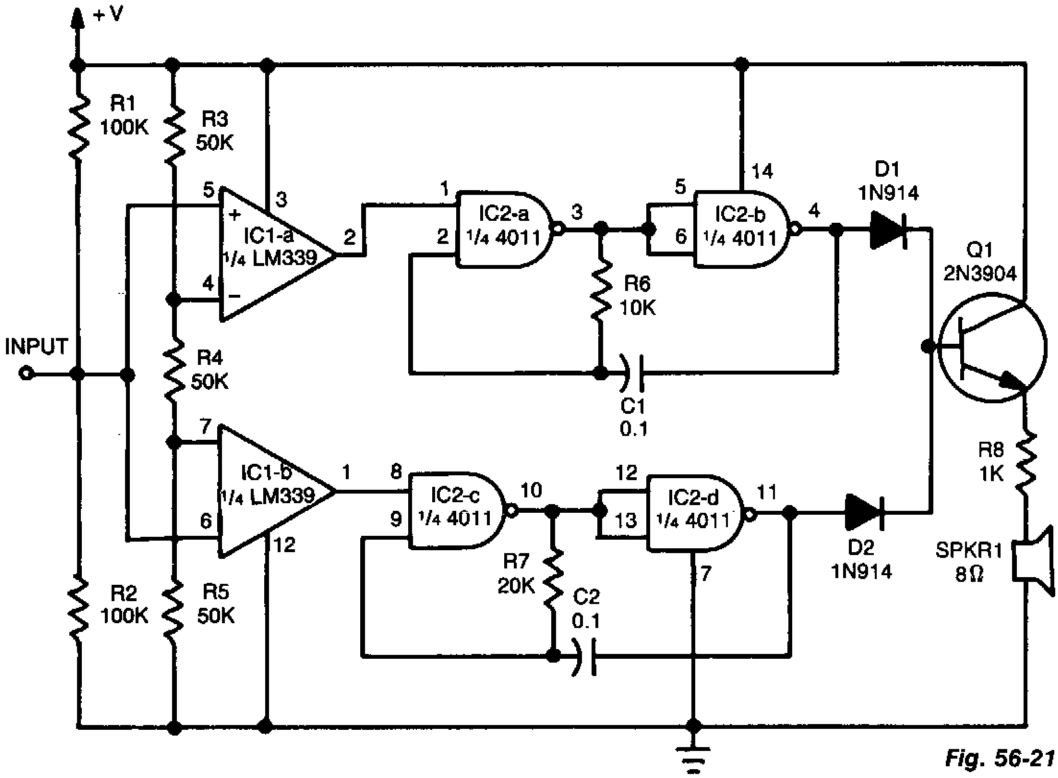
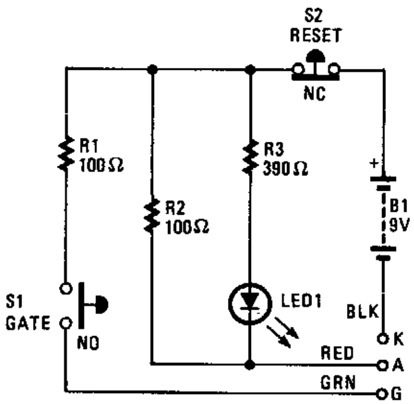


Fig. 56-21

This tester provides an audible indication of the logic level of the signal presented to its input. A logic high is indicated by a high tone, a logic low is indicated by a low tone, and oscillation is indicated by an alternating tone. The input is high impedance, so it will not load down the circuit under test. The tester can be used to troubleshoot TTL or CMOS logic. The input consists of two sections of an LM339 quad comparator. IC1a increases when the input voltage exceeds 67% of the supply voltage. The other comparator increases when the input drops below 33% of the supply.

The tone generators consist of two gated astable multivibrators. The generator built around IC2a and IC2b produces the high tone. The one built around IC2c and IC2d produces the low tone. Two diodes, D1 and D2, isolate the tone-generator outputs. Transistor Q1 is used to drive a low-impedance speaker.

SCR TESTER

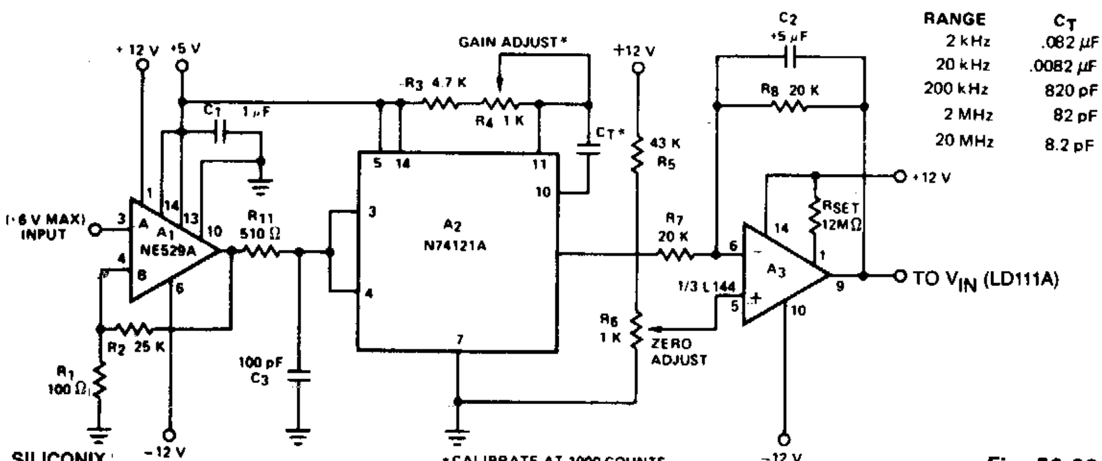


HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 56-22

The DUT's(Device Under Test) cathode, anode, and gate are connected to the unit's K, A, and G terminals, respectively. Pressing switch S1 feeds a gate current to the DUT, which triggers it on. Resistor R1 limits the gate current to the appropriate level. Resistor R3 limits the current through the LED to about 20 mA, which, with the current through R2, results in a latching current of about 110 mA. The LED is used to monitor the latching current. If the DUT is good, once the gate is triggered with S1, the LED will remain lit, indicating that the device is conducting. To end the test, turn off the device by interrupting the latching current flow using switch S2. The LED should turn off and remain off. The preceding procedure will work with SCRs and triacs. To check LEDs and other diodes, connect the anode and cathode leads to the anode and cathode of the diode; LED1 should light. When the leads are reversed, the LED should remain off.

DIGITAL FREQUENCY METER

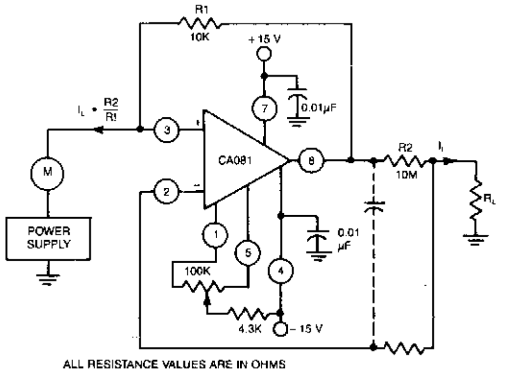


SILICONIX

Fig. 56-23

The circuit converts frequency to voltage by taking the average dc value of the pulses from the 74121 monostable multivibrator. The one shot is triggered by the positive-going ac signal at the input of the 529 comparator. The amplifier acts as a dc filter, and also provides zeroing. This circuit will maintain an accuracy of 2% over 5 decades of range. The input signal to the comparator should be greater than 0.1 V pk-pk, and less than 12 V pk-pk for proper operation.

LOW-CURRENT MEASUREMENT SYSTEM

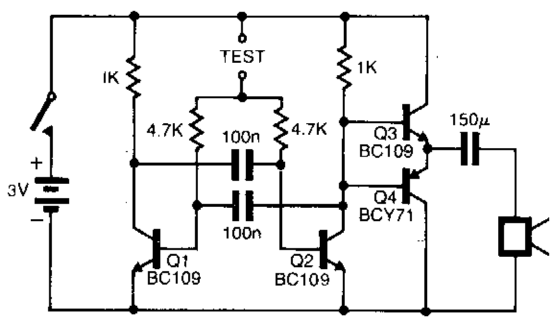


GE/RCA

Fig. 56-24

This circuit uses a CA018 BiMOS op amp. Low current, supplied at input potential as power supply to load resistor R_L, is increased by R2/R1, when load current I_L is monitored by power supply meter M. Thus, if I_L is 100 nA, with values shown, I_L presented to supply will be 100 μA.

SIMPLE CONTINUITY TESTER

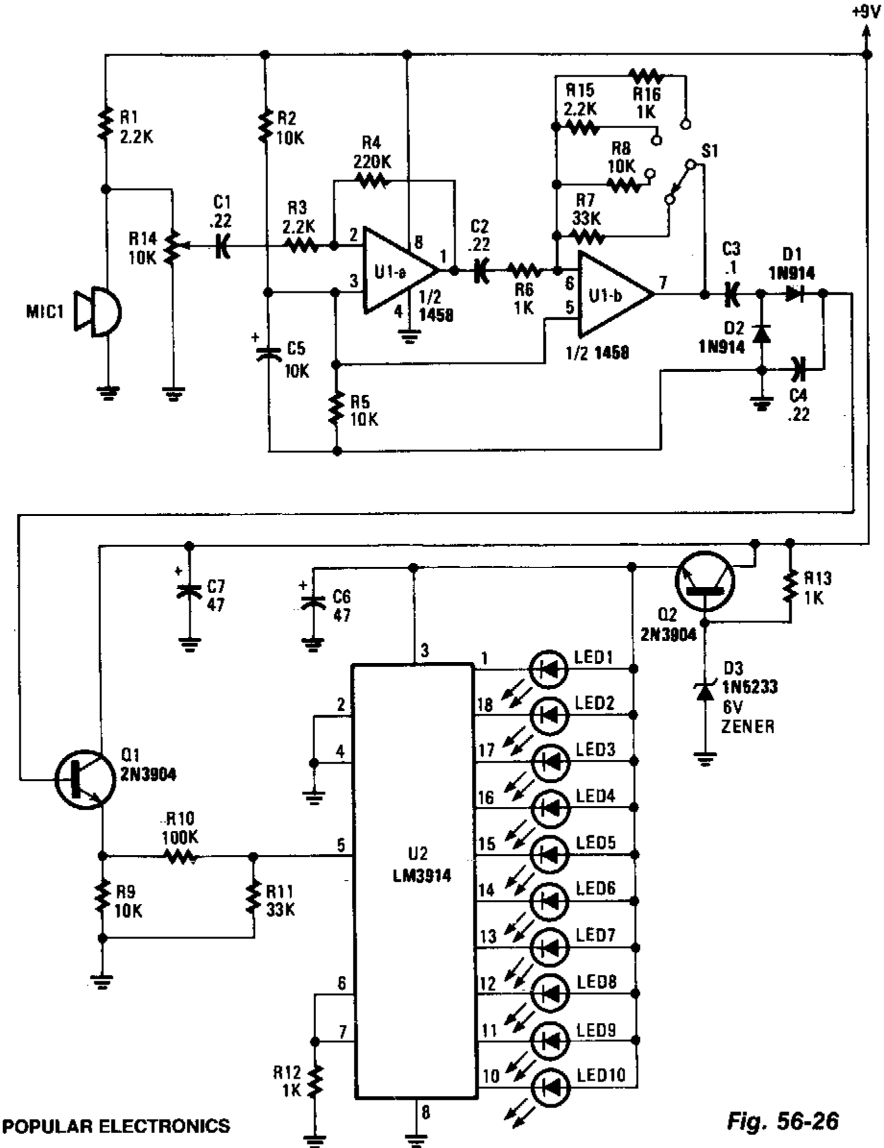


ELECTRONIC ENGINEERING

Fig. 56-25

The pitch of the tone is dependent upon the resistance under test. The tester will respond to resistances of hundreds of kilohms, yet it is possible to distinguish differences of just tens of ohms in low-resistance circuits. Q1 and Q2 form a multivibrator, the frequency of which is influenced by the resistance between the test points. The output stage, Q3 and Q4, will drive a small loudspeaker or a telephone earpiece. The unit is powered by a 3-V battery, and draws very little current when not in use.

SOUND-LEVEL METER



POPULAR ELECTRONICS

Fig. 56-26

Sounds are picked up by MIC1 and fed to the input of the first op amp. The signal is then fed to the input of second op amp U1b, where it is boosted again by a factor of between 1 and 33, depending upon the setting of range switch S1.

With the range switch set in the A position, R6 is 1 KΩ and R7 is 33 KΩ, so that stage has a gain of 33. In the B position, the gain is 10 Ω; in the C position, the gain is 22 Ω; and in the D position the gain is 1 Ω.

As the signal voltage fed to the input of U2 at pin 5 varies, one of ten LEDs will light to correspond with the input-voltage level. At the input's lowest operating level, U2 produces an output at pin 1, causing LED1 to light. The highest input level presented to the input of U2, about 1.2 V, causes LED10 to turn on.

LED PANEL METER

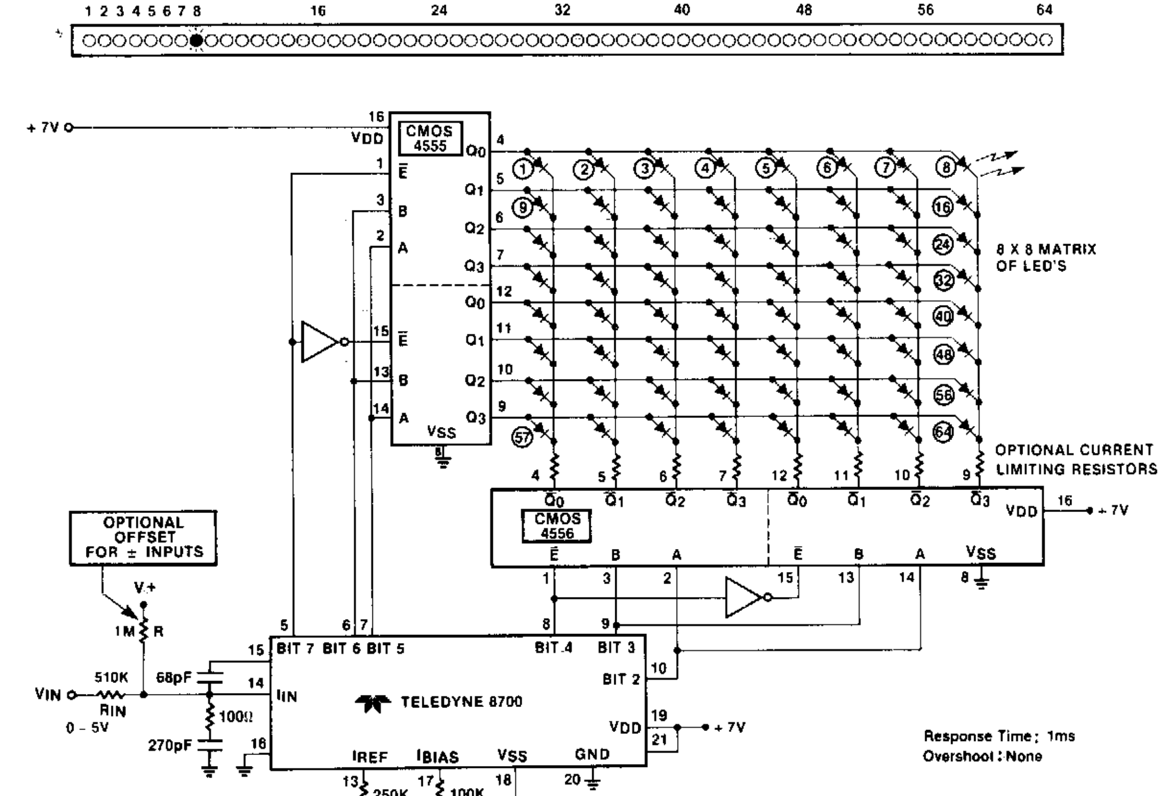
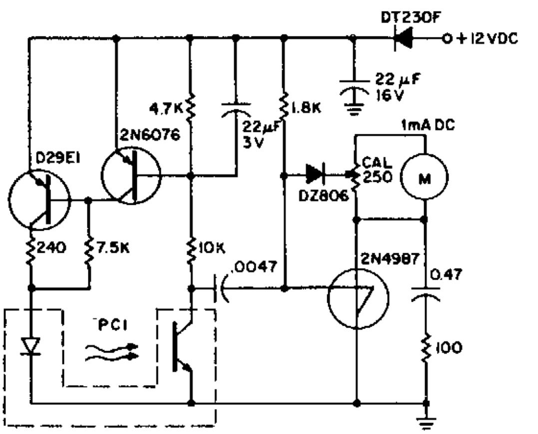


Fig. 56-27

OPTICAL PICK-UP TACHOMETER



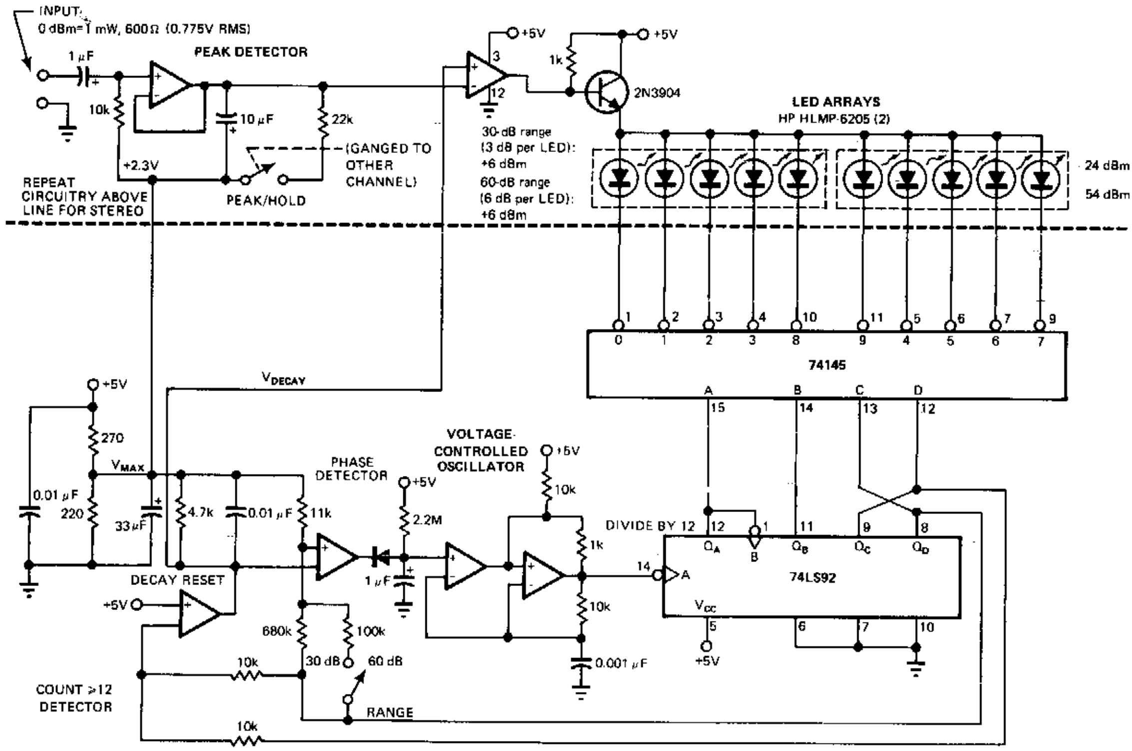
PC-1 AND R { H21A1 FOR TRANSMISSIVE SENSING F5D1 AND L14G3 FOR REFLECTIVE SENSING GFOE1A1 AND GFOD1A1 FOR FIBER OPTIC PROBE

GE

Fig. 56-28

Remote, noncontact, measurement of the speed of rotating objects is the purpose of this simple circuit. Linearity and accuracy are extremely high and normally limited by the milliammeter used and the initial calibration. This circuit is configured to count the leading edge of light pulses and to ignore normal ambient light levels. It is designed for portable operation since the tachometer is not sensitive to supply voltage within the supply voltage tolerance. Full scale at the maximum sensitivity of the calibration resistance is read at about 300 light pulses per second. A digital voltmeter can be used, on the 100-mV full-scale range, in place of the milliammeter. Shunt its input with a 100-Ω resistor in parallel with a 100-μF capacitor. This rc network replaces the filtering supplied by the analog meter.

PEAK-dB METER



EDN

Fig. 56-29

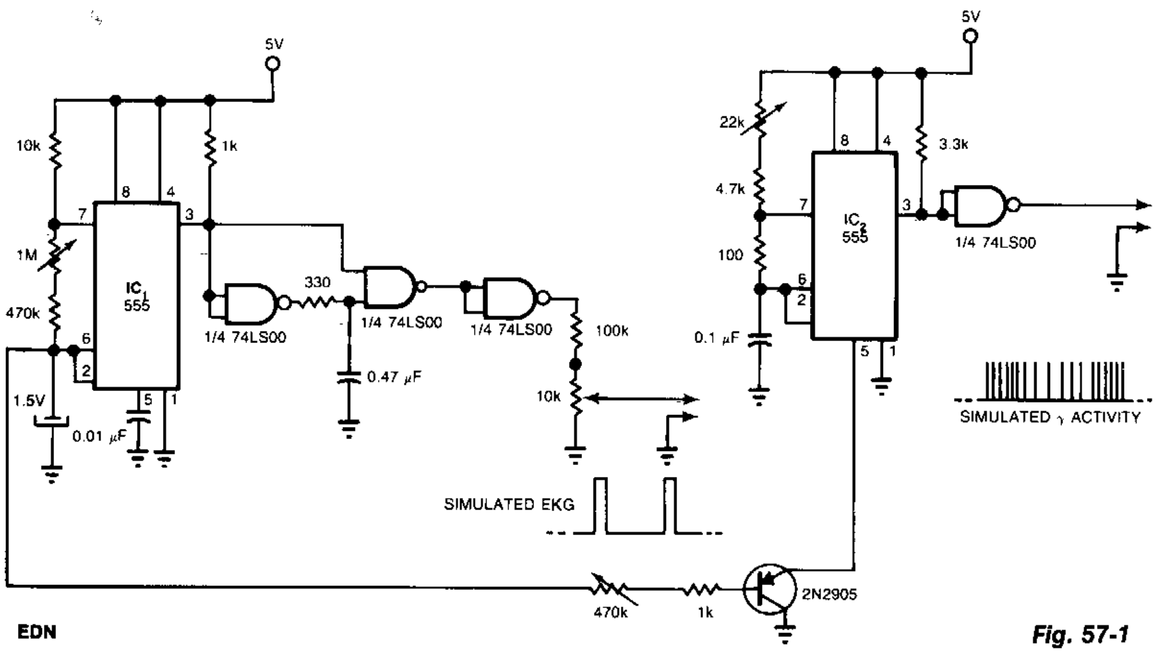
This circuit compares a rectified input, V_{IN}, with a voltage that decays exponentially across a 4.7-KΩ resistor and a 0.01-μF capacitor. Comparing the exponentially decaying voltage with the rectified input provides a peak-level indication that requires no adjustment. A phase-locked loop controls the scan rate so that each LED represents 6 dB in the 30 dB range.

57

Medical Electronics Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

THREE-CHIP EKG SIMULATOR

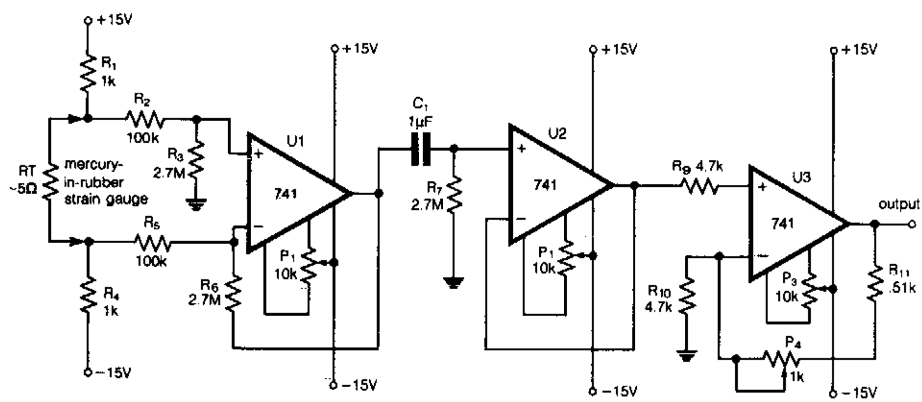


EDN

Fig. 57-1

Two 555s and a quad NAND-gate IC can simulate an electrocardiograph signal and a γ-wave radioisotope signal for applications in nuclear medicine. This circuit synchronizes the radioisotope signal to the EKG signal. You can use the circuit's outputs to test, for example, microprocessor-based software that calculates the left ventricular ejection fraction before you use the software in clinical applications. IC1, a 555 timer, provides a positive-going pulse train that simulates an EKG signal. A 10-KΩ potentiometer provides frequency adjustment. The other 555 timer, configured as a pulse-position modulator, provides the simulated γ-wave activity.

BREATH MONITOR

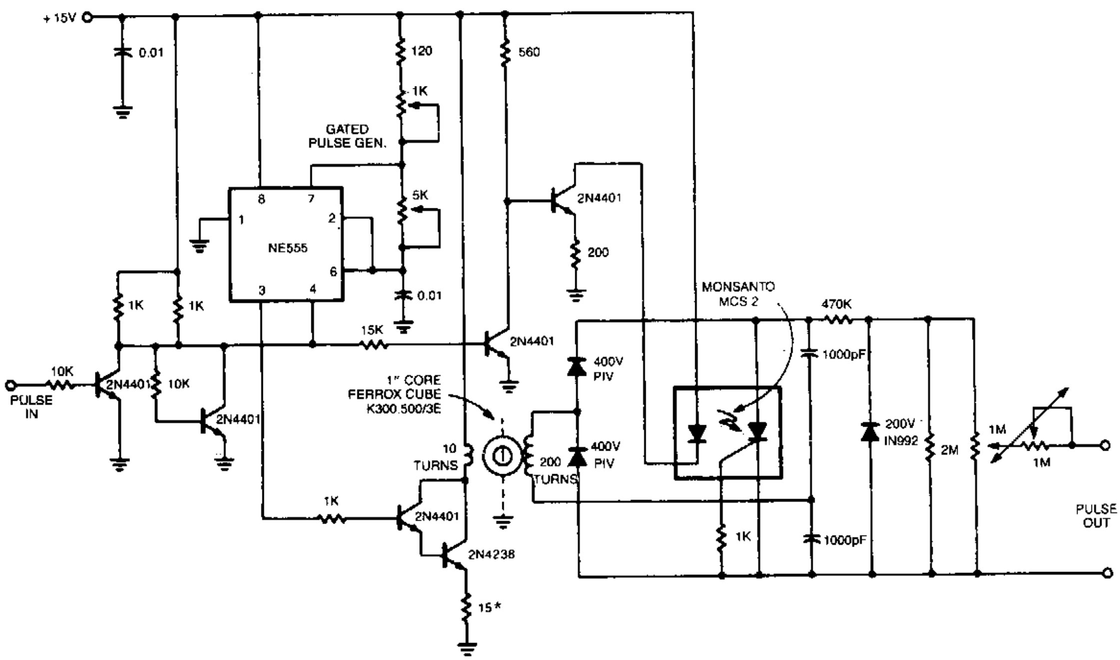


ELECTRONIC ENGINEERING

Fig. 57-2

The mercury-in-rubber strain gauge is the detector of breathing. In the model device, the strain gauge produced by Medimatic, Demark was used. The change in the length of the strain gauge wrapped around the chest during breathing causes a varying electrical resistance of about 0.2 Ω/cm. The constant current passing R1, R_T and R4 gives the constant component on the output of U1 differential amplifier. The change in the resistance of the strain gauge, which results from breathing, produces proportional varying in output voltage of U1 amplifier. The voltage follower, based on U2, separates the output voltage of the filter from the input stage amplifier. U3 works as a noninverter amplifier with regulated gain. This device should be supplied with a stabilized constant voltage of ±15 V.

STIMULUS ISOLATOR



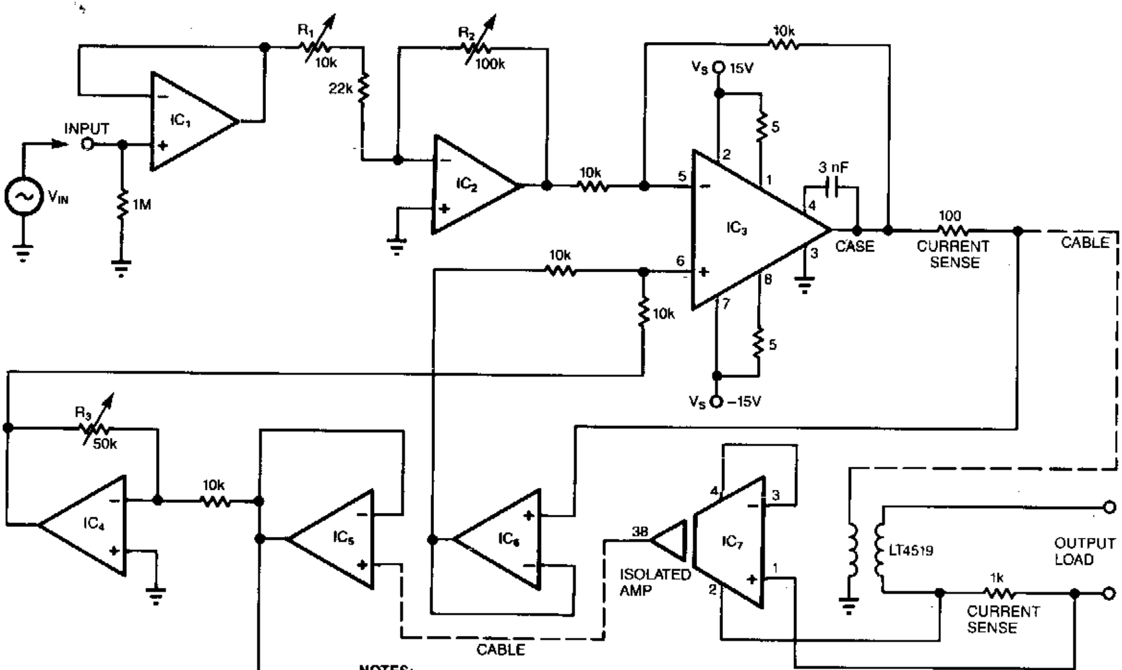
NOTES:
All resistor values in ohms
*Power rating depends on duty cycle from 1/2w for 20-25% duty cycle to 15-20W for 75-90% duty cycle

SIGNETICS

Fig. 57-3

This stimulus isolator uses a photo-SCR and a toroid for shaping pulses of up to 200 V at 200 μA.

CONSTANT-CURRENT STIMULATOR



NOTES:
1. IC3 IS LH0021.
2. IC7 IS AD202JY.
3. IC1, IC2, IC5, AND IC6 ARE EACH ¼TL074CN.
4. LT4519 IS FROM LOUTH TRANSFORMER CO, LOUTH, LINCOLNSHIRE, ENGLAND.

EDN

Fig. 57-4

Most circuits that provide an electrical stimulus for research subjects are constant-voltage designs; this circuit is a constant-current design. Stimulator circuits must be isolated for two reasons: to ensure safety and to minimize interference. Isolated stimulators are essentially two-terminal devices; output currents can flow only between the two output terminals and can at no time flow through any other path, such as the power ground.

The circuit's bandwidth ranges from 50 Hz to 5 kHz when a ± 1 V sinusoidal input drives the circuit. Output loads can range from a short circuit to 100 KΩ and have as much as 0.033 μF of parallel capacitance. The transformer and associated circuitry conveniently connect to the main circuit via a cable. Note: This circuit is not approved for use on human beings.

Op amps, IC1 and IC2, buffer and set the gain of the circuit, respectively. You adjust trimmer R1 so that R2, a 10-turn pot, yields output currents ranging from 0 to 1 mA/V_{IN}. IC3 is a power op amp. Its output drives the primary of a transformer that has a current gain of 0.1, or a voltage gain of 10. Operating from a +15 V supply, the transformer therefore has a voltage compliance of ± 150 V.

The circuit senses not only the current supplied to the transformer but also the current in the transformer's secondary. IC7, a fully isolated, medical-grade amplifier, provides the isolated feedback signal because the op amp has its own built-in isolation transformer. Trimmer R3 sets the feedback gain precisely at 27 KΩ nominal.

58

Metronome

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DOWNBEAT-EMPHASIZED ELECTRONIC METRONOME

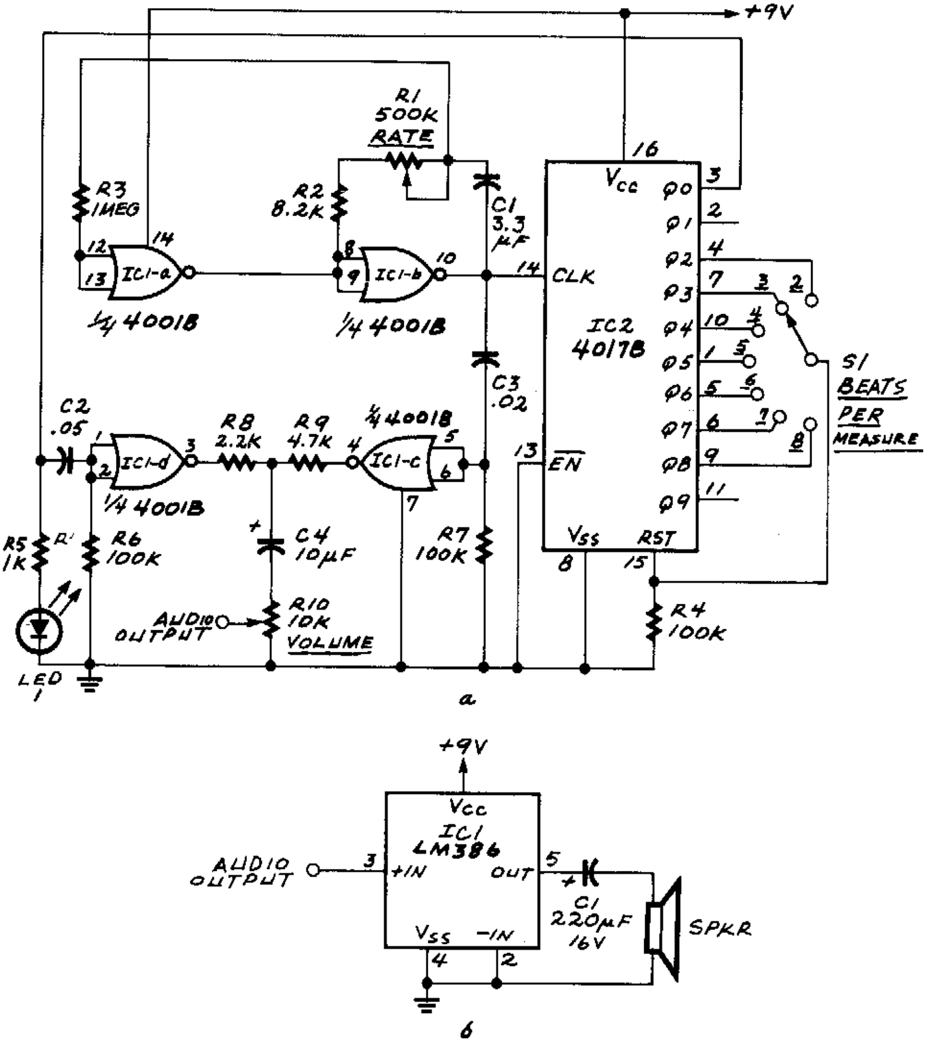


Fig. 58-1

IC1a and IC1b form an astable multivibrator. The astable's signal is fed to IC1c, also to the clock input of IC2, a 4017B decade counter. That IC's Q0 through Q9 outputs become high one at a time for each successive clock pulse received at pin 14. Switch S1 feeds one of those outputs to the 4017B's reset input. Whenever the selected output becomes high, the 4017B restarts its counting cycle; that determines the number of beats per measure. The network composed of C2 and R6 sharpens the downbeat pulse, and the network composed of C3 and R7 sharpens the free-running pulses. By making C2 larger than C3, the downbeat receives greater emphasis.

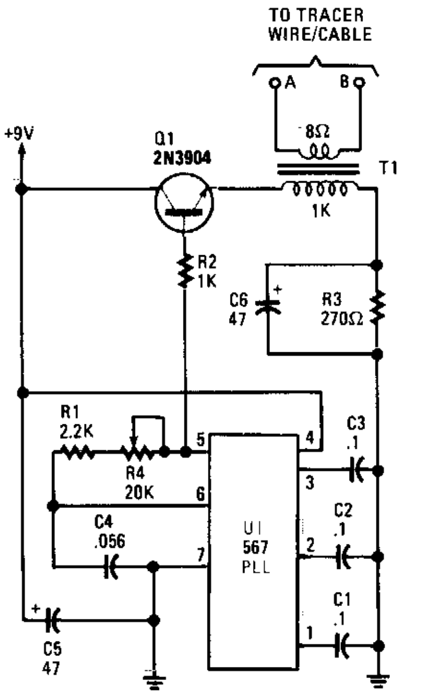
59

Miscellaneous Treasures

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

- Closed-Loop Tracer
- Pulse Amplitude Discriminator
- Tracer Receiver
- Central Image Canceller
- Bug Tracer
- Breath Alert Alcohol Tester
- Automatic Electronic Music
- Positive Input/Negative Output Charge Pump
- Acid Rain Monitor
- Simple Low-Cost Rf Switch
- Burst Power Control
- Flame Ignitor
- Bar Code Scanner
- 50-MHz Trigger
- Air-Motion Detector
- Bug Detector
- Door Opener

CLOSED-LOOP TRACER



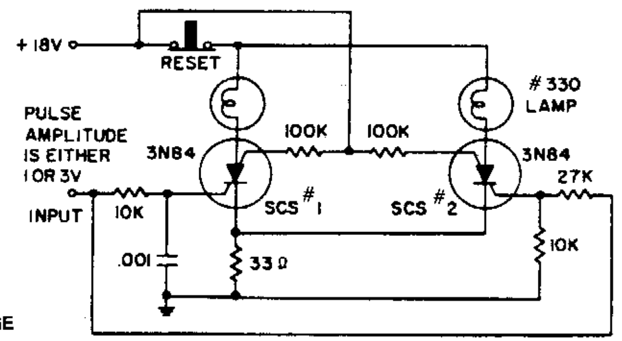
POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

Fig. 59-1

The tracer, consisting of both a transmitter and receiver, is designed to follow a closed-loop wire or cable system. It follows an induced voltage path created by feeding a low-current, audio-frequency signal through the cable. When the pick-up coils come within close proximity of the current-carrying cable, a small voltage is generated in each coil, and that induced voltage is then processed by the receiver's circuitry.

The circuit is built around a 567 phase-locked loop (PLL) configured as a variable-frequency, audio-generator circuit, which is designed to produce a square-wave output at pin 5. Potentiometer R4 allows the oscillator to be easily tuned to the receiver's frequency. Transistor Q1 isolates the oscillator from the load and matches the impedance of the primary of T1. Resistor R3 limits current flow through Q1. The low-impedance secondary of T1 supplies the cable drive signal.

PULSE AMPLITUDE DISCRIMINATOR

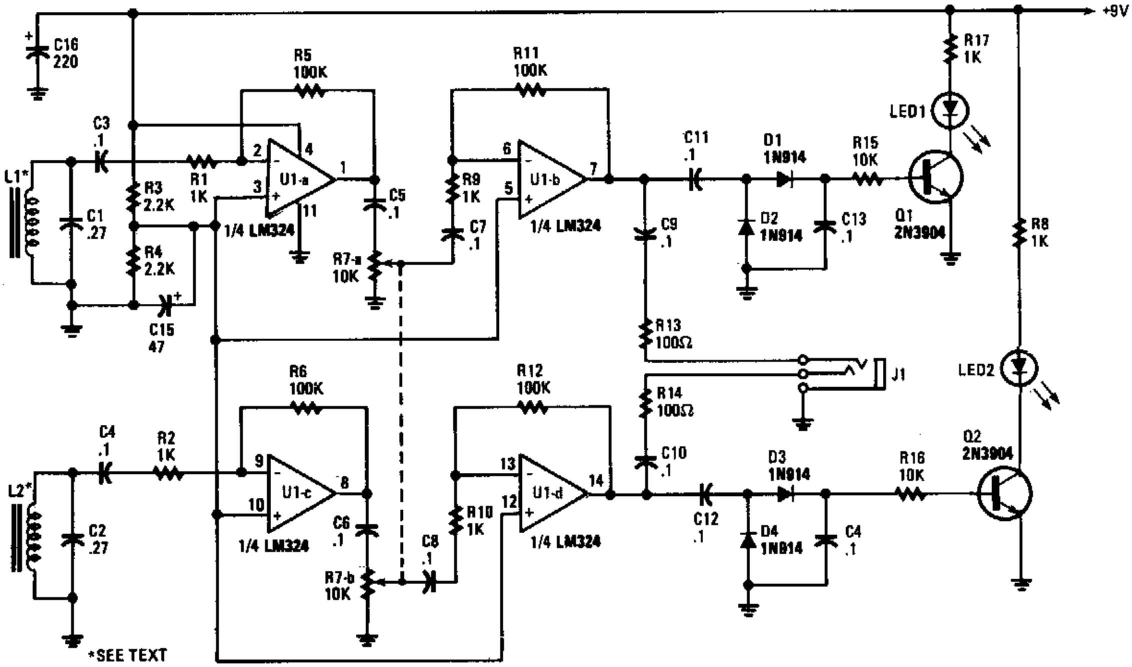


GE

Fig. 59-2

A 1-V amplitude pulse triggers SCS1, but has insufficient amplitude to trigger SCS2. A 3-V input pulse is delayed in reaching SCS1 by the 10-KΩ and .001-μF integrating network. Instead, it triggers SCS2, then raises the common emitter voltage to prevent SCS1 from triggering. The 100-KΩ resistors suppress the rate effect.

TRACER RECEIVER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

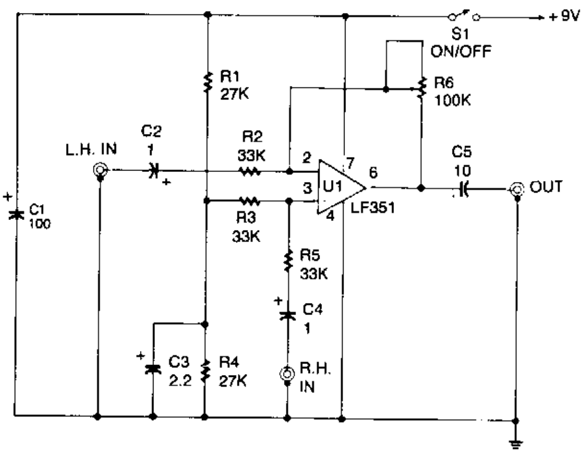
Fig. 59-3

The tracer receiver is a stereo audio amplifier/detector circuit operating near 1 kHz. Inductors L1 and L2—hand-wound coils, consisting of 200 turns of #26 wire on 2-inch ferrite cores— are tuned to the operating frequency of the amplifier/detector. The received signal strength of each individual receiver is indicated by an LED. The audio output of the receiver is fed to a stereo headphone. That dual-receiver scheme helps in locating and tracking the hidden wire or cable by giving a directional output that indicates the cable's path.

The 1-kHz signal is picked up by L1 and coupled to the input of op amp U1a, which provides a gain of about 100 dB. The output of op amp U1a is fed through volume-control potentiometer R7 to the input of U1b, which magnifies the already amplified signal 100 times more. That puts the maximum gain of the receiver at about 10,000 dB. The output of U1b follows two paths: in the first path, the signal is couple through C9 and R13 to J1, and is used to drive one half of a stereo headphone.

In the other path, the signal is fed through a voltage doubling/detector circuit—consisting of D1, D2, C11, and C13—that converts the amplified 1-kHz signal to the dc voltage that's used to drive Q1. When Q1 is turned on, LED1 lights, indicating a received signal.

CENTRAL IMAGE CANCELLER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

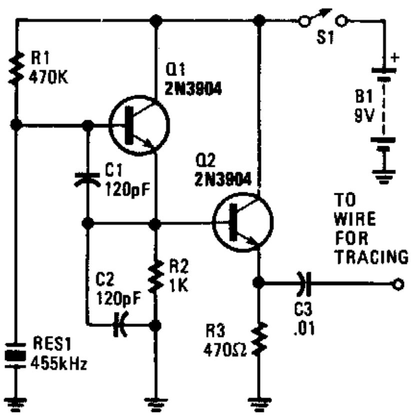
Fig. 59-4

The circuit allows you to eliminate the vocal portion of an audio signal, while leaving the instrumental portion. The circuit mixes two channels that must be 180° out of phase, so the signals that form the center-stereo image is canceled out. Those signals usually appear in phase. Resistor R3 biases the noninverting input of U1 from a center tap formed by resistors R1 and R4, and capacitor C3. Resistor R4, capacitor C3, and potentiometer R6 form a negative-feedback circuit that establishes the closed-loop voltage gain of U1 at unity. The signal is inverted between the input and output.

Signals applied to the right input are coupled to the noninverting input of U1 through C4 and attenuating resistor R5. Resistors R3 and R5 make up a 6 dB attenuator, so once again, there is unity voltage gain between the input and the output. However, the right input signal is not inverted.

Therefore, a signal appearing at both inputs is phased out by the circuit and will not appear at the output. Even if the two input signals are at slightly different levels because of different source impedances, you can still adjust for full cancellation by carefully tweaking R6.

BUG TRACER



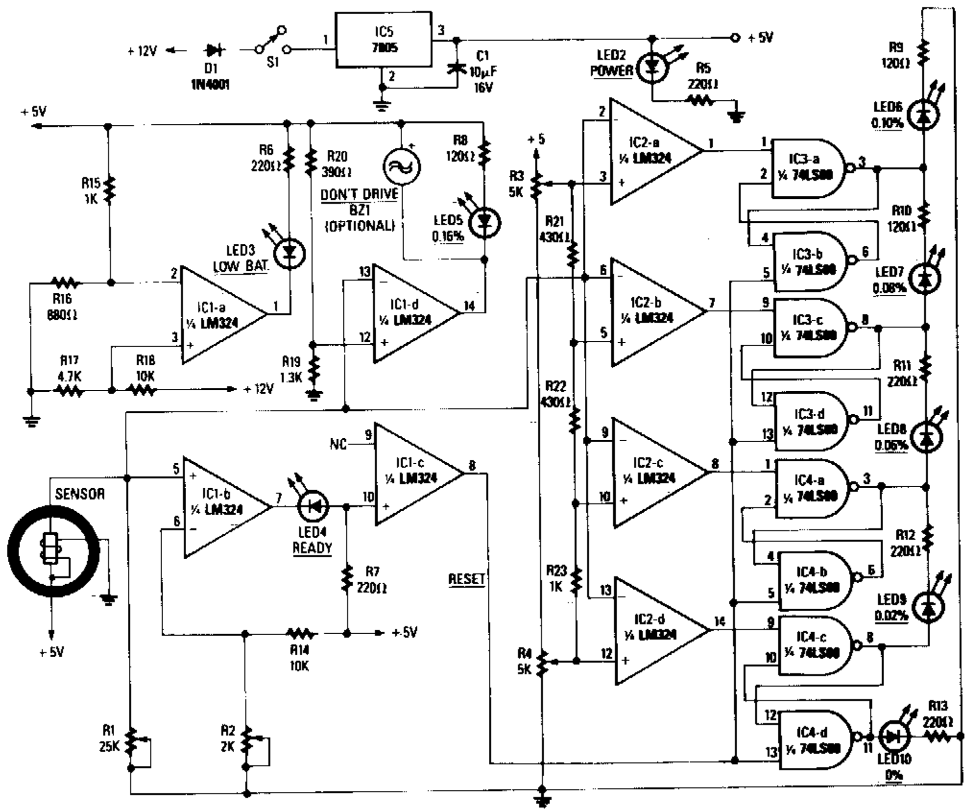
POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

Fig. 59-5

The bug tracer is made up of a simple rf-injector circuit consisting of Q1 and Q2, and a pocket-size, AM broadcast receiver. The two-transistor rf-injector circuit supplies a constant rf signal to one end of a cable. Then the AM receiver is used as a detector, allowing you to trace the wire to its source.

Transistor Q1, along with piezoelectric ceramic resonator RES1, make up a simple rf oscillator that operates either at or near the AM-radio, 455-kHz, i-f frequency. That means that the second or third harmonic signal can easily be picked up by the receiver. Transistor Q2 is connected to an emitter-follower circuit to protect the oscillator from output loading; that helps to stabilize the output frequency and signal level.

BREATH ALERT ALCOHOL TESTER



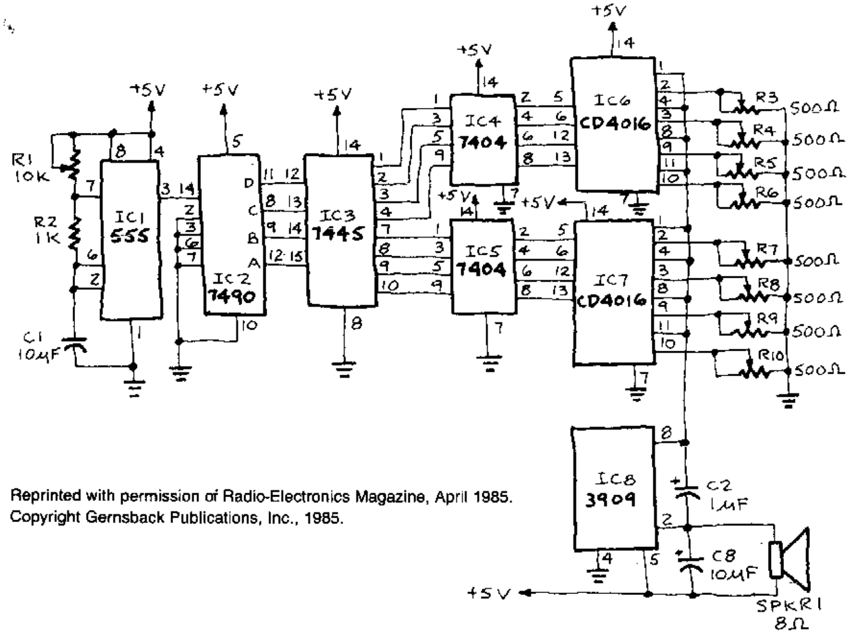
Reprinted with permission from Radio-Electronics Magazine, October 1988. Copyright Gernsback Publications, Inc., 1988. ***Fig. 59-6***

When power is applied to the circuit, the heater coil in the sensor is energized by the 5-V output of IC5, a 7805 voltage regulator. Breathing into the sensor with alcohol on your breath will lower the sensor's resistance; consequently, the input voltage to the detector circuit, will change. The detector circuit consists of quad op amp, IC2 and its associated circuitry. All sections of the detector circuit are calibrated via R3 and R4, and the inputs to each section are controlled by the voltage-divider network R21 through R23. As each section is triggered, the outputs decrease, and sample-and-hold circuits, IC3 and IC4, will latch onto the highest input value and drive the appropriate LED. The different colored LEDs represent alcohol levels from 0 to 0.16%.

If the level of alcohol is above the legal limit, or 0.16%, part of another quad op amp, IC1d, will turn on both the optional buzzer and LED5. That is an indication of a high level of alcohol present in your blood, and you definitely should not drive.

After a test is taken,the sensor takes a few seconds to ready itself for another test. When the sensor is ready, its input to IC1b, adjusted via R2 to a threshold of 0.5 V, causes LED4 (ready) to light. That, in turn, causes IC1c to reset the rest of the circuitry. The last section of IC1 is biased via R15 and R16, and used to indicate a low-battery condition—when the battery voltage drops below 6.8 V—which could result in an inaccurate breath test.

ELECTRONIC MUSIC

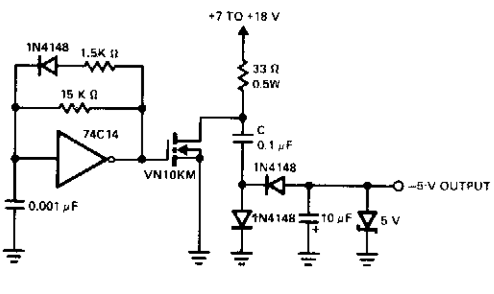


Reprinted with permission of Radio-Electronics Magazine, April 1985.

Fig. 59-7

IC1, a 555 timer, is set up as an astable multivibrator to produce the signal that triggers IC2, a 7490 decade counter. That IC, in turn, produces a BCD output that is fed to IC3, a 7445 BCD-to-decimal decoder/driver. IC3's output is inverted by two hex inverters, IC4 and IC5. The outputs of IC4 and IC5 are inputted to control pins on IC6 and IC7, CD4016 CMOS quad bilateral switches. As those switches open and close, different resistances (as set by potentiometers R3 through R10) are inserted into the sound-generating circuit made from IC8. The frequency at the outputs of IC6 and IC7 are adjusted to various rates, using potentiometers R3 through R10, to produce the desired tones. Capacitors can be placed in series with the potentiometers to produce a sloping sound instead of a straight tone. The negative-going output signals of IC6 and IC7 are fed through a common bus to pin 8 of IC8.

POSITIVE INPUT/NEGATIVE OUTPUT CHARGE PUMP



SILICONIX

Fig. 59-8

A charge pump is a simple means of generating a low-power voltage supply of opposite polarity from the main supply. The 74C14 IC is a self-oscillating driver for the MOSFET power switch. It produces a pulse width of 6.5 µs at a repetition frequency of 100 kHz. When the MOSFET device is off, capacitor C is charged to the positive supply. When the power through the MOSFET switches on, C delivers a negative voltage through the series diode to the output. The zener serves as a dissipative regulator. Because the MOSFET switches fast, operation at high frequencies allows the capacitors in the system to be small.

ACID RAIN MONITOR

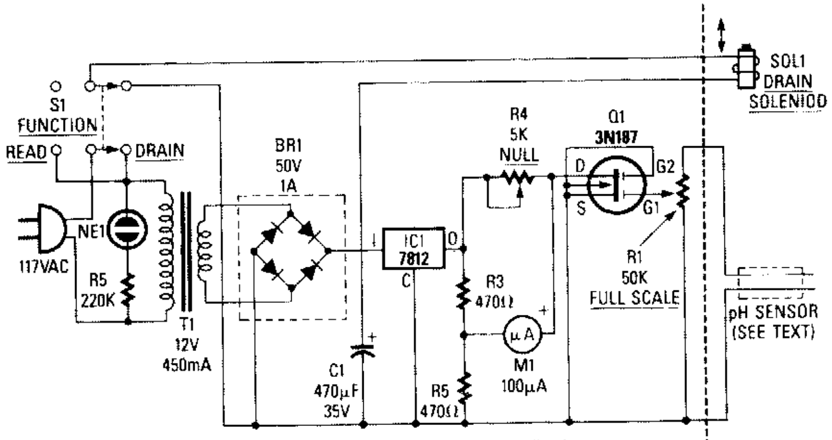
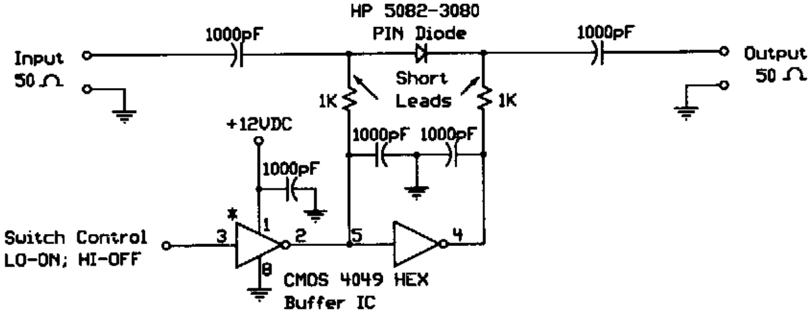


Fig. 59-9

A simple bridge rectifier and a 12-V regulator powers the MOSFET sensing circuit. The unregulated output of the bridge rectifier operates the drain solenoid via switch S1. The sensor itself is built from two electrodes: one made of copper, the other of lead. In combination with the liquid trapped by the sensor, the electrodes form a miniature lead/acid cell whose output is amplified by MOSFET Q1. The maximum output produced by our prototype cell was about 50 μA.

MOSFET Q1 serves as the fourth leg of a Wheatstone bridge. When sensed acidity causes the sensor to generate a voltage, Q1 turns on slightly, so its drain-to-source resistance decreases. That resistance variation causes an imbalance in the bridge, and that imbalance is indicated by meter M1.

SIMPLE LOW-COST RF SWITCH



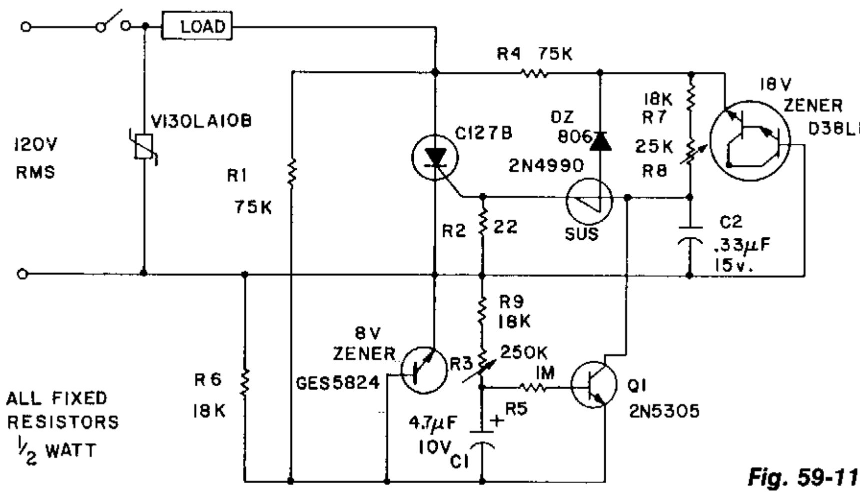
T. Harris, RF Design, July 1989, p. 53.

* This gate is needed only if CMOS driver is not present

Fig. 59-10

When the digital logic level at the control input is low, the PIN diode is forward-biased by the CMOS gates. The two 1-KΩ bias resistors limit this current to the PIN diode's safe forward current limit. In this state, the switch is on. When the control input is high, the diode is reverse-biased and the switch is off. This switch is well-suited for electronically steered antenna arrays, multiple path switching, and other applications requiring small, low-cost rf switches. This particular design was used in a four-pole rotary switch for a Doppler-shift radio direction-finder operating at 144 MHz.

BURST POWER CONTROL



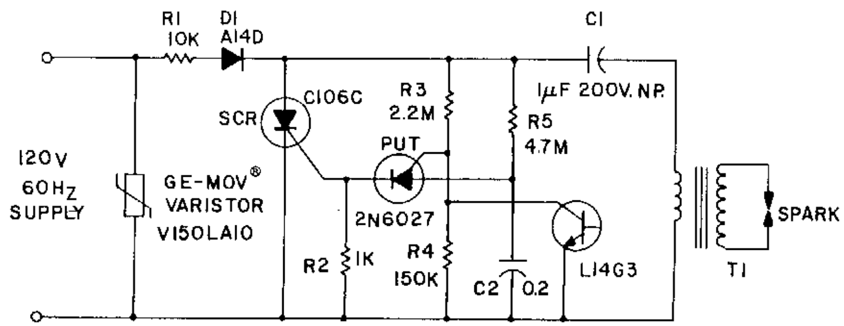
GE

Fig. 59-11

Industrial applications sometimes require that power be only briefly applied to a load following the closure of a switch, such as a microswitch or foot switch. The load could be a heating element for use in sealing plastic bags, a dc motor that is indexed or stepped with each application of power, or a dc solenoid which is to be energized for a brief time—for example, a staple gun.

The phase angle at which the SCR triggers on is determined by the charging rate of C2 through R7 and potentiometer R8. When the breakover voltage of the gate triggering device, a silicon unilateral switch, is reached, the SCR turns on. The average dc voltage to the load is determined by the setting of R8.

FLAME IGNITOR



T1 Primary – 8T No. 16 Litz
Secondary – 800T No. 28
Core 1/2" dia. x 2" Ferrite Rod

For 220V Operation:
GE-MOV® V275LA15
SCR C106D
D1 A14N

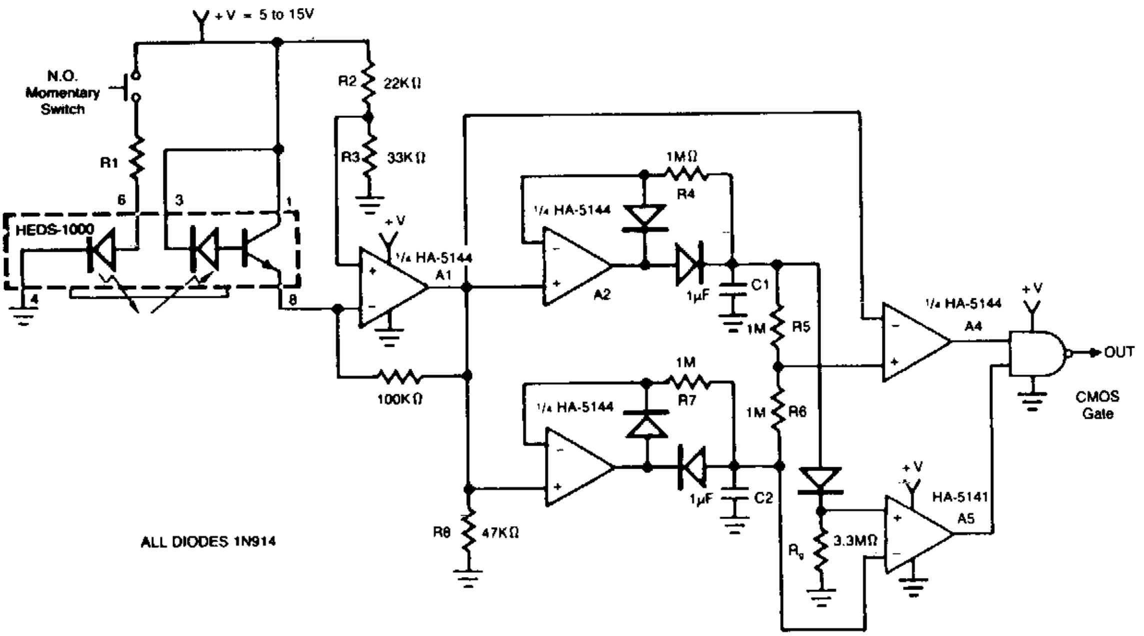
C1 .33 µF 400V NP
R3 4.7 M
R5 10. M

GE

Fig. 59-12

The spark developed by the circuit is suitable for a gas ignitor. Capacitor C1 is charged through R1 and D1 toward peak line voltage. C2 is simultaneously being charged at a slower rate through R5. When the charge on C2 is sufficient to trigger the PUT, the SCR is triggered on, providing a rapid discharge path for C1 through the transformer primary. The SCR is triggered about 20 times per second with the component values shown. The L14G3 serves as a flame sensor. When ignition is achieved and sensed by the photodetector, the low V_{CE} prevents further SCR triggering.

BAR-CODE SCANNER

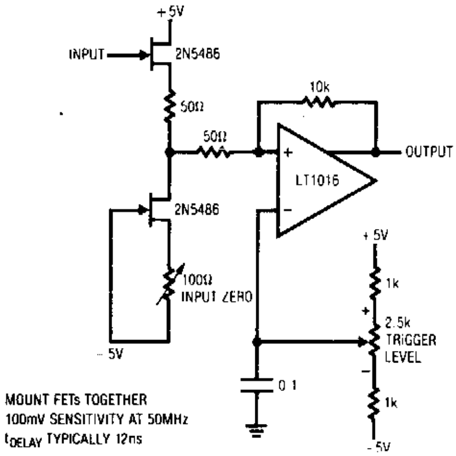


HARRIS

Fig. 59-13

The circuit illustrates a method of interfacing a HEDS-1000 emitter-detector pair with a HA-5144 for use as a bar-code scanner circuit. The HA-5144 is used as an amplifier system which converts the bar and space widths of the printed bar code into a pulse-width modulated digital signal. Amplifier A1 is used to amplify the current output of the detector. The output of A1 is passed to two precision peak-detector circuits which detect the positive and negative peaks of the received signal. Amplifier A4 is used as a comparator whose reference is maintained at the midpoint of the peak-to-peak signal by resistors R5 and R6. This provides a more accurate edge detection and less ambiguity in bar width. Amplifier A5 is used as an optional noise gate which only allows data to pass through the gate when the peak-to-peak modulation signal is larger than one diode drop. This circuit is operated by a single supply voltage with low-power consumption which makes it ideal for battery-operated data entry systems.

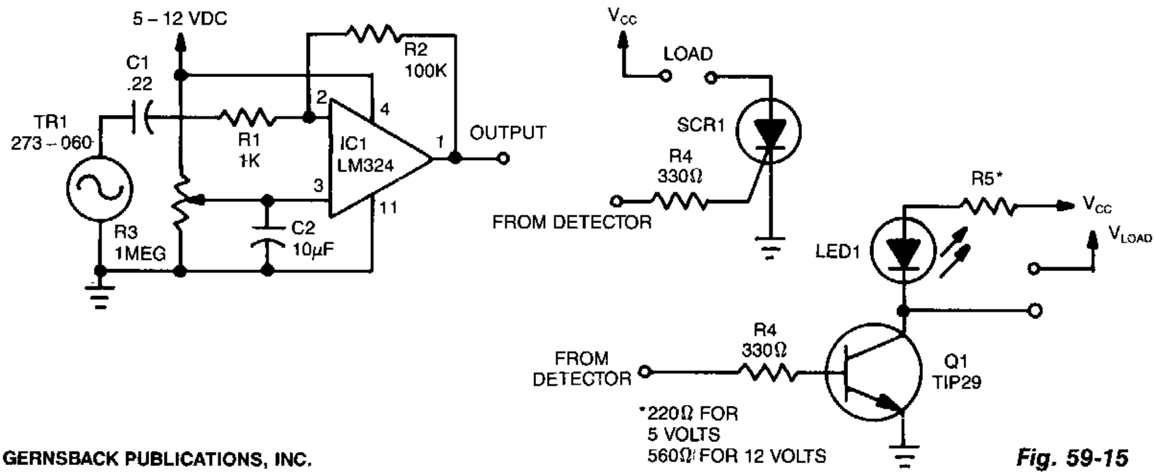
50-MHz TRIGGER



LINEAR TECHNOLOGY CORP. *Fig. 59-14*

This has a stable trigger 100 mV sensitivity at 50 MHz. The FETs comprise a simple high-speed buffer and the LT1016 compares the buffer's output to the potential at the trigger level potentiometer, which can be of either polarity. The 10-KΩ resistor provides hysteresis, eliminating "chattering" caused by noisy input signals. To calibrate this circuit, ground the input and adjust the input zero control for 0 V at Q2's drain terminal.

AIR-MOTION DETECTOR



GERNSBACK PUBLICATIONS, INC. *Fig. 59-15*

When a current of air hits the piezo element, a small signal is generated and is fed through C1 and R1 to inverting input pin 2 of one section of the LM324. That causes output pin 1 to increase. Resistor R3 is used to adjust the sensitivity of the detector. The circuit can be set so high as to detect the wave of a hand or so low that blowing on the element as hard as you can will produce no output. Resistor R2 is used to adjust the level of the output voltage at pin 1.

The detector circuit can be used in various control applications. For example, an SCR can be used to control 117-Vac loads as shown. Also, an npn transistor, such as a TIP29, can be used to control loads as shown.

BUG DETECTOR

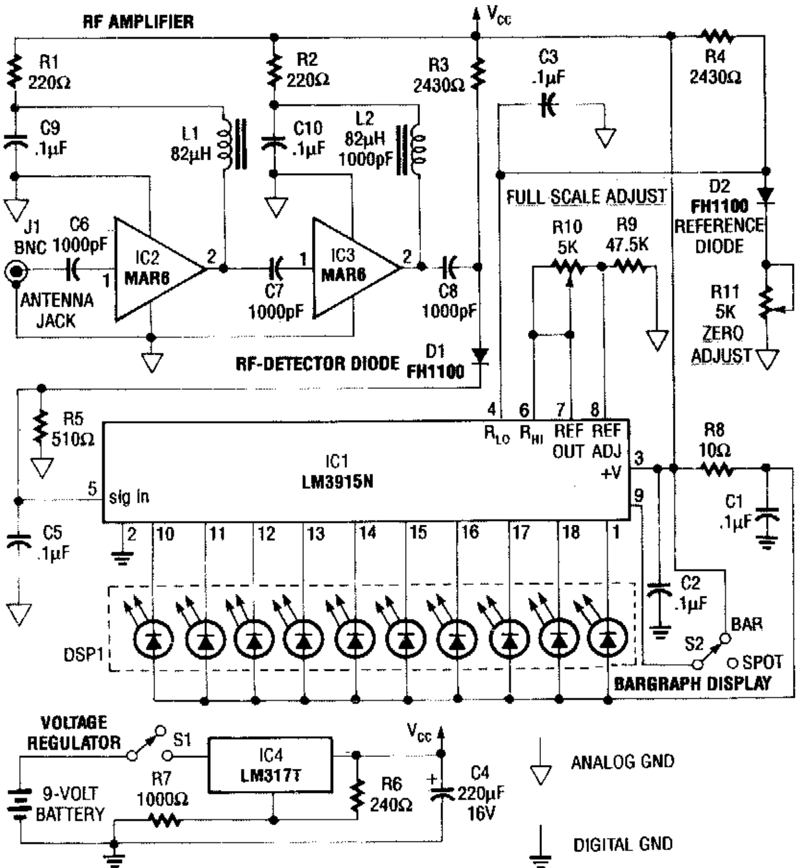
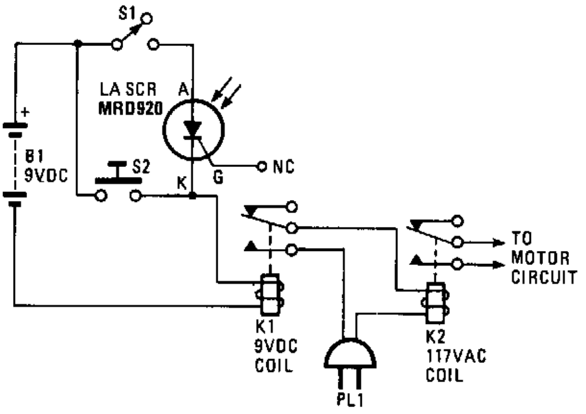


Fig. 59-16

This rf detector can locate low-power transmitters (bugs) that are hidden from sight. It can sense the presence of a 1-mW transmitter at 20 feet, which is sensitive enough to detect the tiniest bug. As you bring the rf detector closer to the bug, more and more segments of its LED bar-graph display light, which aids in direction finding.

The front end has a two-stage wideband rf amplifier, and a forward-biased hot-carrier diode for a detector. After detection, the signal is filtered and fed to IC1, an LM3915N bar-graph driver having a logarithmic output. Each successive LED segment represents a 3-dB step.

DOOR OPENER



HANDS-ON ELECTRONICS

Fig. 59-17

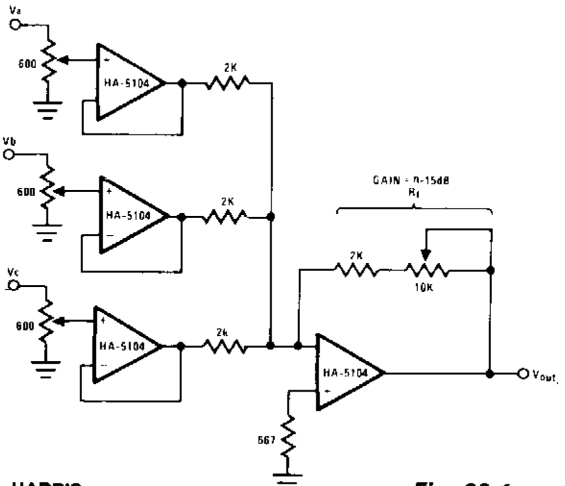
The door opener derives its power from a 9-V battery. A momentary-contact switch, S2, is provided in the event that manual opening and closing is required. Relay K1 is a 9-V type and relay K2 is a 117-Vac latching-type, which automatically latches with the first burst of current and opens on the second burst. The gate lead of the LASCR is not used; a light source triggers the LASCR unit into conduction, causing current to flow in the coil of the relay. That, in turn, causes K1's contacts to close, thereby energizing K2 (closing its contacts), and operating the garage door motor.

60

Mixers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SIGNAL COMBINER

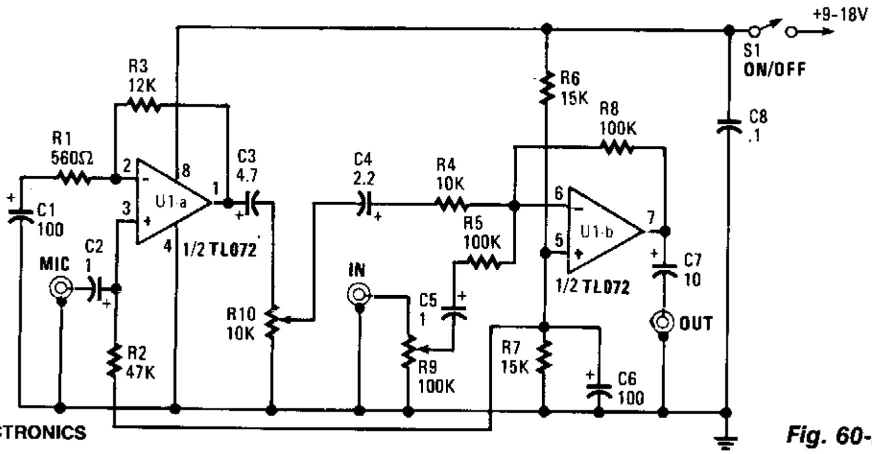


HARRIS

Fig. 60-1

This circuit uses buffer stages to prevent channel crosstalk back through the mixer resistor network. The potentiometers used for each stage allow for convenient signal strength adjustment, while maintaining input impedance matching at the 600-Ω audio standard. The feedback resistor R_f will permit the output signal gain to be as high as 15 dB.

SIMPLE MIXER



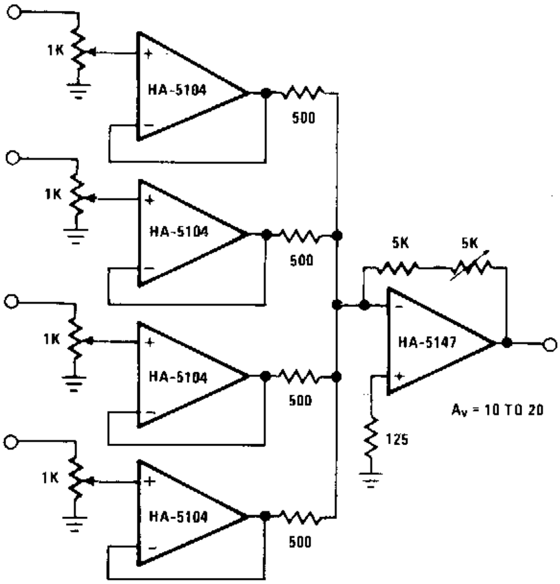
POPULAR ELECTRONICS

Fig. 60-2

This mixer is built around a TL072 dual BiFET op amp with a JFET input stage, and can be powered from a single-ended 9- to 18-V power supply. The microphone input is capacitively coupled to the noninverting input of U1a.

Resistors R1 and R3 set the voltage gain at about 26 dB and serve as a negative feedback network for U1a. Capacitors C1 through C3 are dc-blocking capacitors. Most high-impedance microphones have outputs of a few mV. Often, a preamp stage just isn't enough, so the microphone signal is given a boost of about 20 db in the mixer. The noninverting input of U1b is biased to half the supply voltage by R6, R7, and C6. Resistors R5 and R8 make up the negative-feedback network and set the voltage gain of U1b at unity. Capacitor C5 is for dc blocking at this input.

INPUT-BUFFERED MIXER

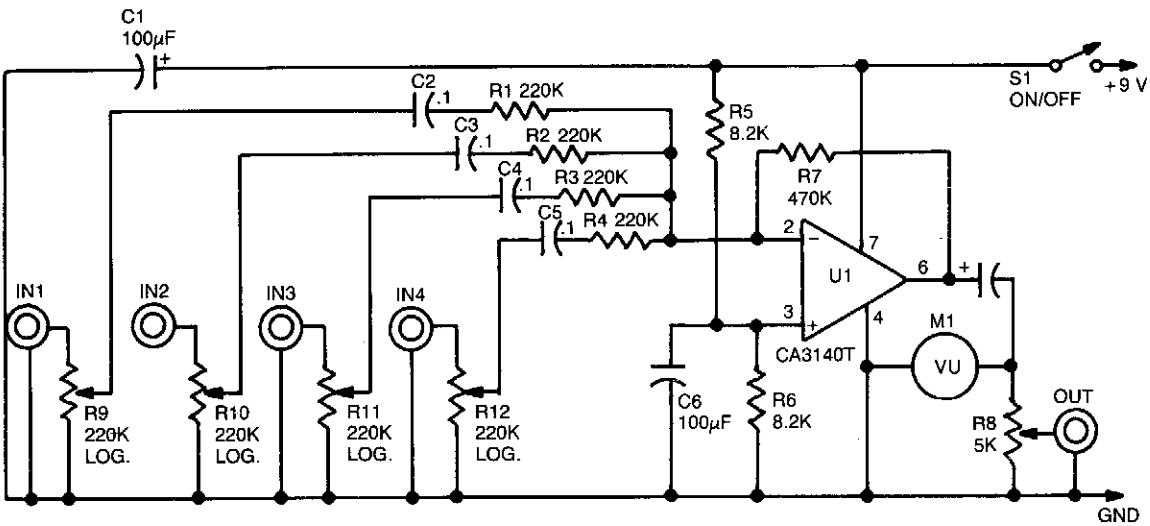


HARRIS

Fig. 60-3

A high signal-to-noise ratio is important in signal construction and combination. The HA-5147 aids in lowering overall system noise and thereby raises system sensitivity. The signal combination circuit incorporates input buffering with several other features to form a relatively efficient mixer stage with a minimum of channel crosstalk. The potentiometer used for each channel allows for both variable input levels and a constant impedance for the driving source. The buffers serve mainly to prevent reverse crosstalk back through the resistor network. This buffering allows for the combination of varying strength signals without reverse contamination. The gain of the final stage is set at a minimum of 10 and can be adjusted to as much as 20. This allows a great amount of flexibility when combining a vast array of input signals.

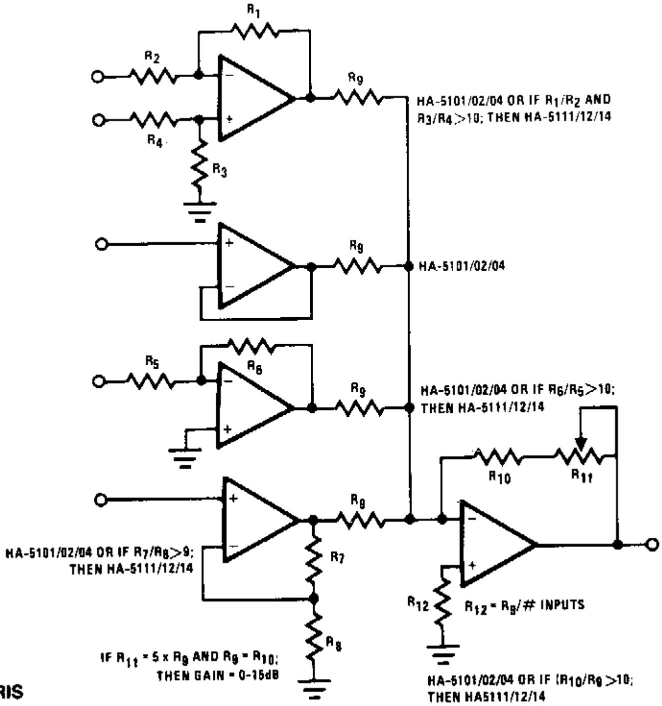
FOUR-CHANNEL MIXER



POPULAR ELECTRONICS

Fig. 60-4

UNIVERSAL MIXER STAGE



HARRIS

Fig. 60-5

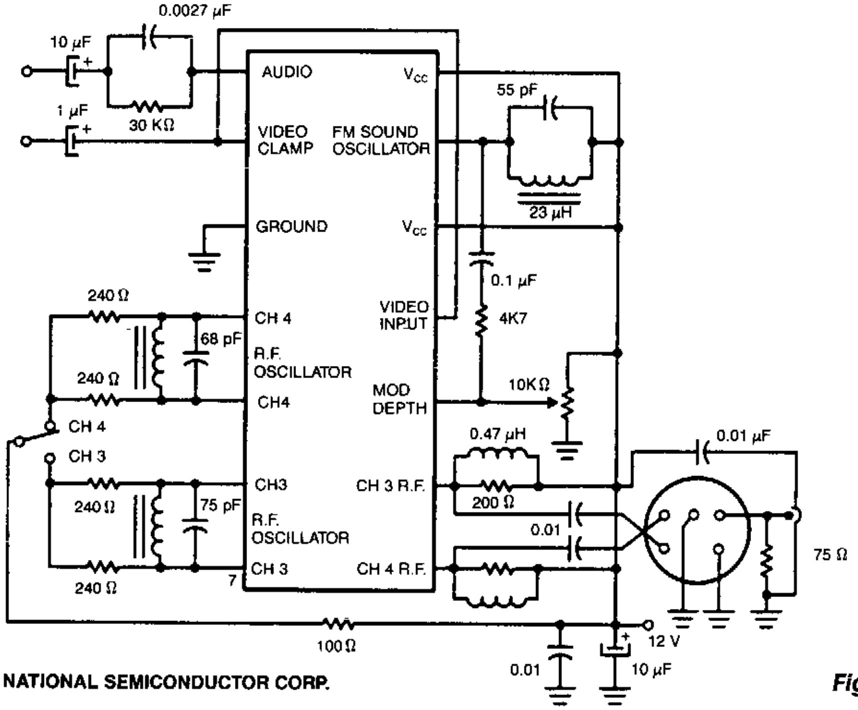
This circuit illustrates some possible buffer combinations. These include a differential input stage, a voltage follower as well as both noninverting and inverting stages. The allowable resistor ratios and recommended device types are also included. One restriction applies to this type of mixer network in which R_g is greater than 2.4 KΩ. This limits the worst case output current for each of the input buffers to less than 10 mA.

61

Modulators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

RF MODULATOR

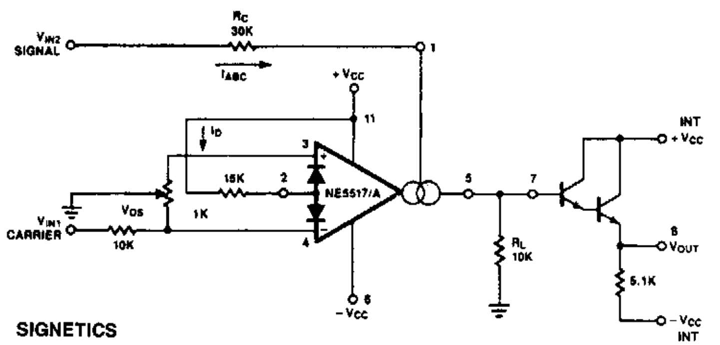


NATIONAL SEMICONDUCTOR CORP.

Fig. 61-1

Two IC rf modulators convert a suitable baseband video and audio signal up to a low VHF modulated carrier—Channel 2 through 6 in the U.S., and 1 through 3 in Japan—the LM1889 and the LM2889. Both ICs are identical regarding the rf modulation function, including pin outs, and can provide either of two rf carriers with dc switch selection of the desired carrier frequency. The LM1889 includes a crystal-controlled chroma subcarrier oscillator and balanced modulator for encoding R through Y or U and V color difference signals. A sound intercarrier frequency lc oscillator is modulated using an external varactor diode. The LM2889 replaces the chroma subcarrier function of the LM1889 with a video dc restoration clamp and an internal frequency-modulated sound intercarrier oscillator.

MODULATOR

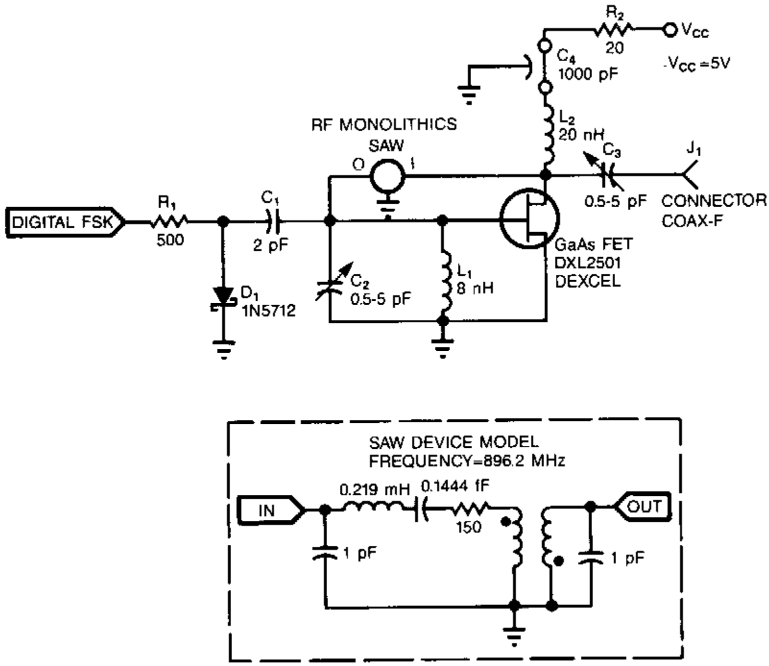


SIGNETICS

Fig. 61-2

Because the transconductance of an operational transconductance amplifier is directly proportional to I_{ABC}, the amplification of a signal can be controlled easily. The output current is the product from transconductance X input voltage. The circuit is effective up to approximately 200 kHz. Modulation of 99% is easy to achieve.

SAW OSCILLATOR MODULATOR



ELECTRONIC DESIGN

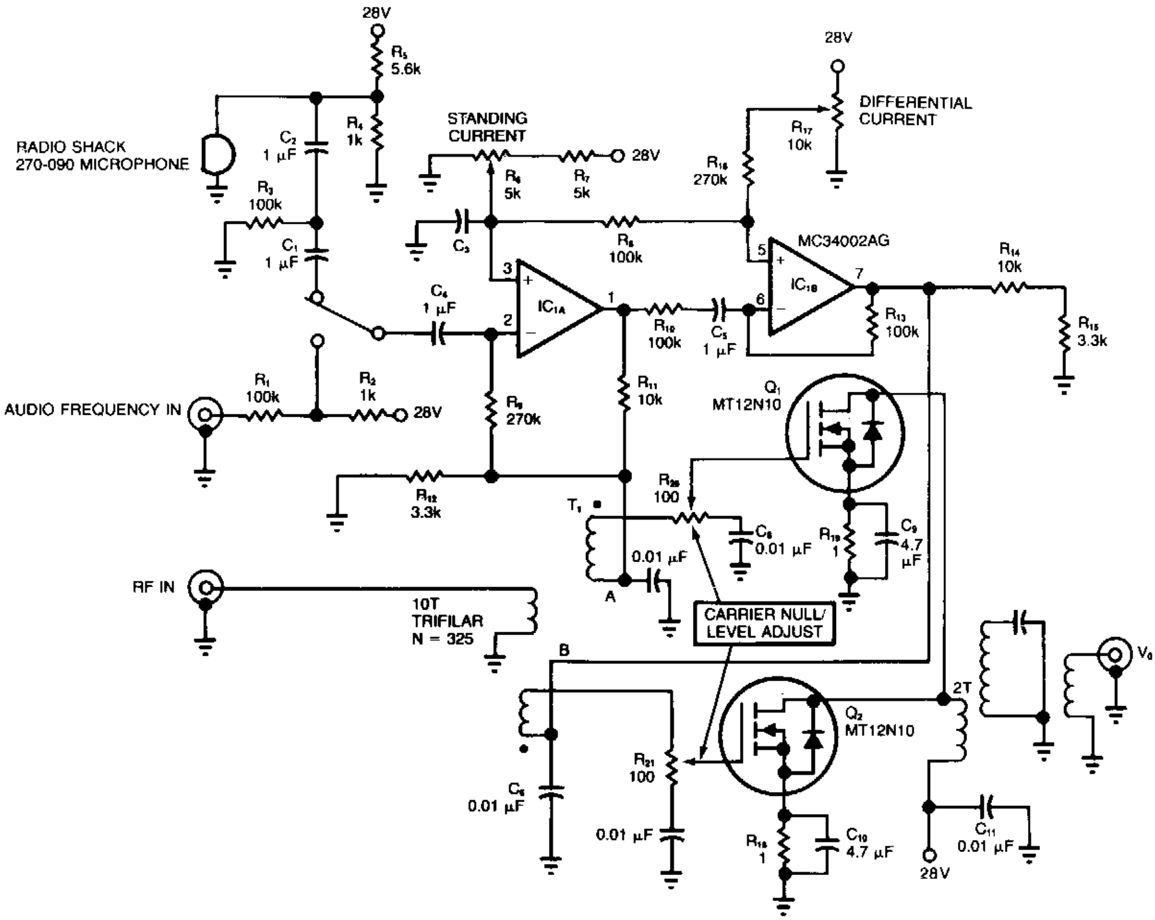
Fig. 61-3

Adding a diode, a resistor, and a capacitor to the SAW (surface-acoustic-wave) oscillator allows you to use the oscillator in FSK (frequency-shift-keying) applications. D1, R1, and C1 form a simple diode switch in which D1 shunts C1 to ground. When the digital FSK input to R1 is low, D1 is off, and the small junction capacitance of D1 couples C1 to ground. A high FSK signal causes current to flow through R1 and D1. D1's dynamic impedance is small when it is in forward conduction. Therefore, C1 sees a lower impedance path to ground. Thus, the FSK input effectively switches C1 in and out of the oscillator's circuit.

When C1 is in the circuit—digital FSK is high—it pulls the frequency of the circuit slightly lower because of the additional phase shift C1 introduces at the GaGs FET gate terminal (available from: Dexcel, Div. of Gould, Santa Clara, CA). The SAW device (available from: RF Monolithics, Dallas, TX) restricts the amount of frequency shifting—usually less than 20 ppm for a high-Q SAW device.

The oscillator shown produces a center frequency of 896.2 MHz with an FSK deviation of 17 kHz when you drive the FSK input with a 0 to 5 V signal. The frequency also depends on L1 and C2.

RF MODULATOR

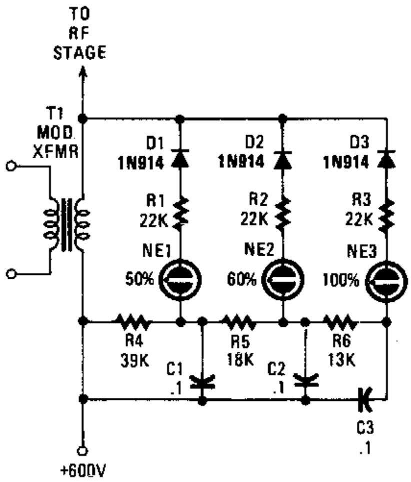


EDN

Fig. 61-4

Because power MOSFETs offer high power gain at both audio and radio frequencies, they are useful in many areas of radio-circuit design. For rf applications, a MOSFET's large safe operating area, V_{DS} vs I_D, protects it against damage from reflected rf energy. As a modulator, a MOSFET's transfer linearity aids fidelity. In the suppressed-carrier modulator, an rf signal is applied to the primary of transformer T1, whose secondaries provide equal-amplitude, opposite-phase rf drive signals to output FETs Q1 and Q2. Output V_o is zero when no audio-frequency signals are present, because the opposite-phase rf signals from Q1 and Q2 cancel. When audio-frequency signals appear at nodes A and B, you obtain a modulated rf output (V_o). Source resistors R18 and R19 improve the dc stability and low-frequency gain. A phase inverter, based on the dual op amp IC1A and IC1B, generates the out-of-phase, equal-amplitude, audio-frequency modulation signals.

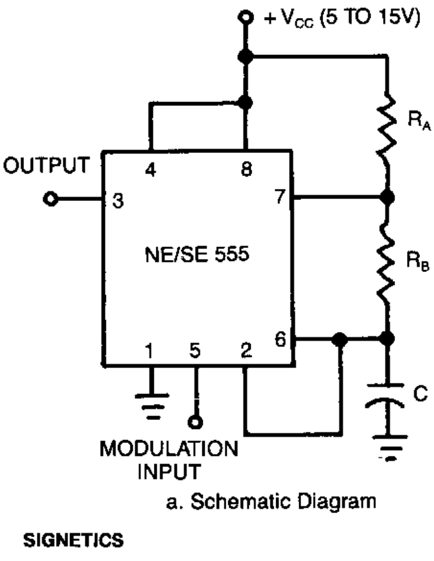
MODULATION MONITOR



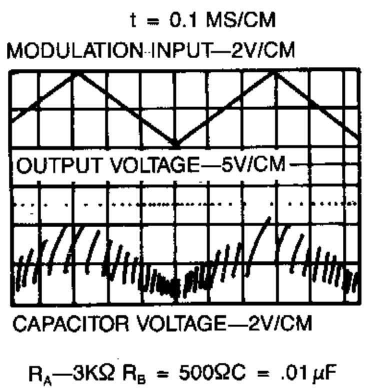
POPULAR ELECTRONICS *Fig. 61-5*

Switching diodes are used to fire the neon lamps when negative-peak modulation hits 50, 60, and 100%. To use the circuit, keep an eye on the lamps. You should attempt to keep the 50% lamp firing all the time, the 60% lamp should be on as much as possible, but try to prevent the 100% lamp from lighting.

PULSE-POSITION MODULATOR



a. Schematic Diagram

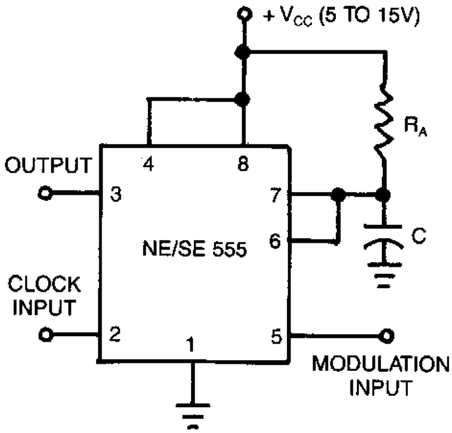


b. Expected Waveform

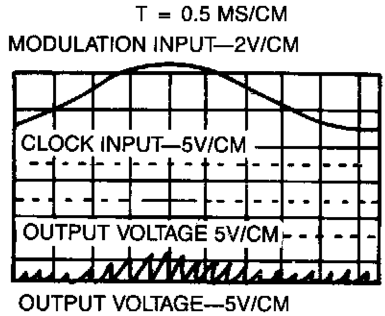
SIGNETICS *Fig. 61-6*

This application uses the timer connected for astable (free-running) operation, with a modulating signal again applied to the control voltage terminal. The pulse position varies with the modulating signal, since the threshold voltage and the time delay is varied.

PULSE-WIDTH MODULATOR



a. Device Schematic



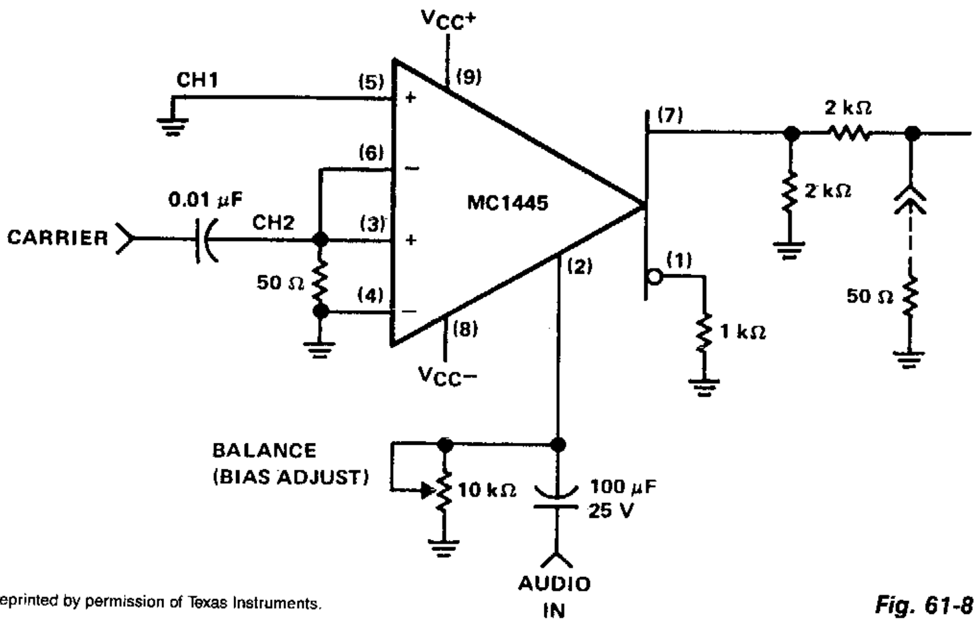
b. Expected Waveforms

SIGNETICS

Fig. 61-7

In this application, the timer is connected in the monostable mode. The circuit is triggered with a continuous pulse train and the threshold voltage is modulated by the signal applied to the control voltage terminal at pin 5. This has the effect of modulating the pulse width as the control voltage varies. The figure shows the actual waveform generated with this circuit.

BALANCED MODULATOR

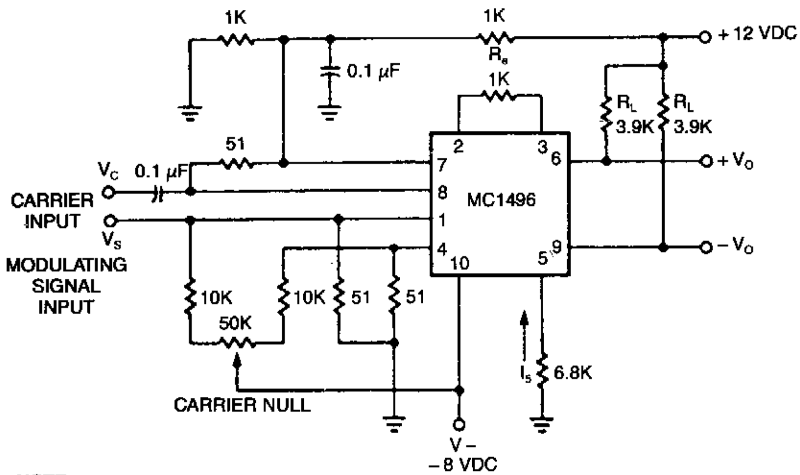


Reprinted by permission of Texas Instruments.

Fig. 61-8

When the carrier level is adequate to switch the cross-coupled pair of differential amplifiers, the modulation signal, which has been applied to the gate, will be switched at the carrier rate, between the collector loads. When switching occurs, it will result in the modulation being multiplied by a symmetrical switching function. If the modulation gate remains in the linear region, only the first harmonic will be present. To balance the MC1445 modulator, equal gain must be achieved in the two separate channels. To remain in the linear region, the modulation input must be restricted to approximately 200 mV pk-pk. Because the gate bias point is sensitive to the amount of carrier suppression, a high-resolution, 10-turn potentiometer should be used.

DOUBLE-SIDEBAND SUPPRESSED-CARRIER MODULATOR



NOTE:
All resistor values are in ohms

SIGNETICS

Fig. 61-9

The basic current allows no carrier to be present in the output. By adding offsets to the carrier differential pairs, controlled amounts of carrier appear at the output. The amplitude becomes a function of the modulation signal—AM modulation.

62

Motor-Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PWM SERVO AMPLIFIER

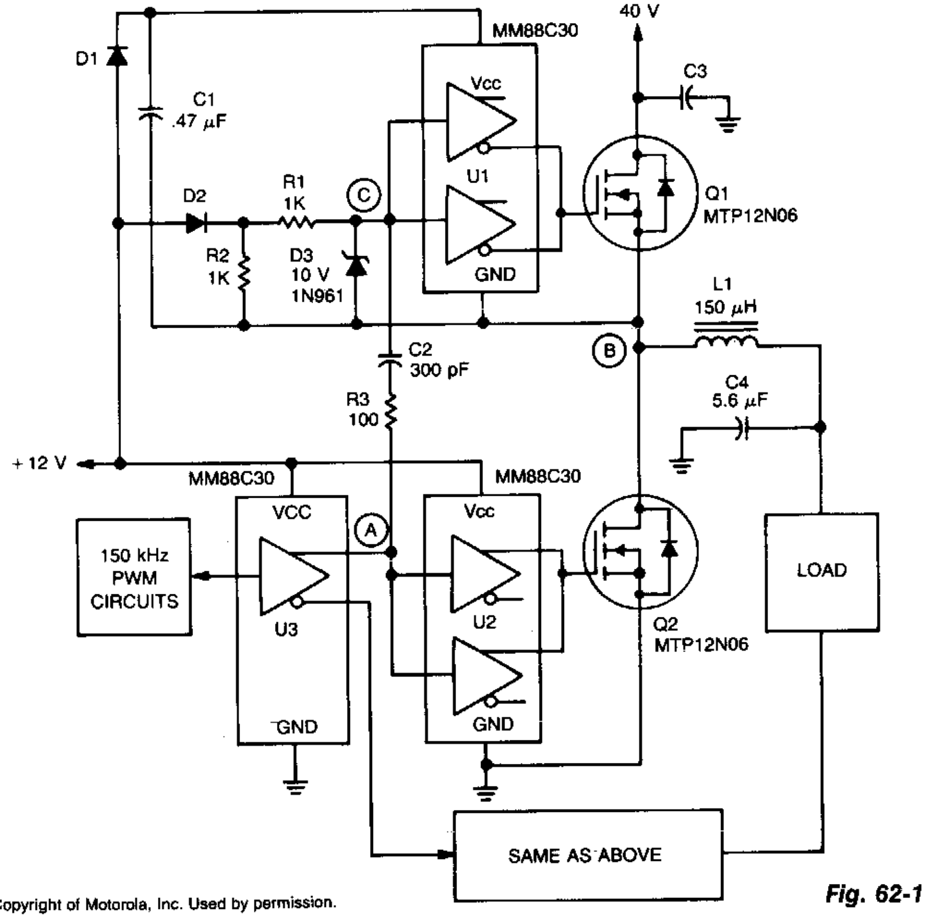


Fig. 62-1

A major feature of the PWM servo amplifier is elimination of a pulse transformer. A 150 kHz pulse-width modulated signal is applied to U3, with its complementary outputs applied to identical circuits to drive the load. When point A increases, Q2 is on and point B is at ground potential. The V_{CC} for U1 is maintained through D1, and Q1 is held off by D2. When point A decreases, Q2 turns off, point C is pulled low by C2, which turns Q1 on. The time constant for R1, R3, and C2 can hold Q1 on just long enough to allow the voltage at point B to start rising. As point B rises, it charges C2 by forward biasing D3, maintaining point C low with respect to U1, and keeping Q1 turned on.

With point B at 40 V, D2 is off and point C is held low by R1 and R2, and V_{CC} for U1 is maintained by the charge on C1. When point A increases again, Q2 again turns on, C2 pushes point C high, and turns Q1 off long enough to allow the voltage at point B to start falling. C2 is now discharged by reverse-biased D3, which keeps point C high with respect to U1, and keeps Q1 off. Once point B reaches ground potential, D1 again turns on, recharging C1, and maintaining V_{CC} to U1. D2 also turns on and keeps Q1 off.

PWM SPEED CONTROL AND ENERGY-RECOVERING BRAKE

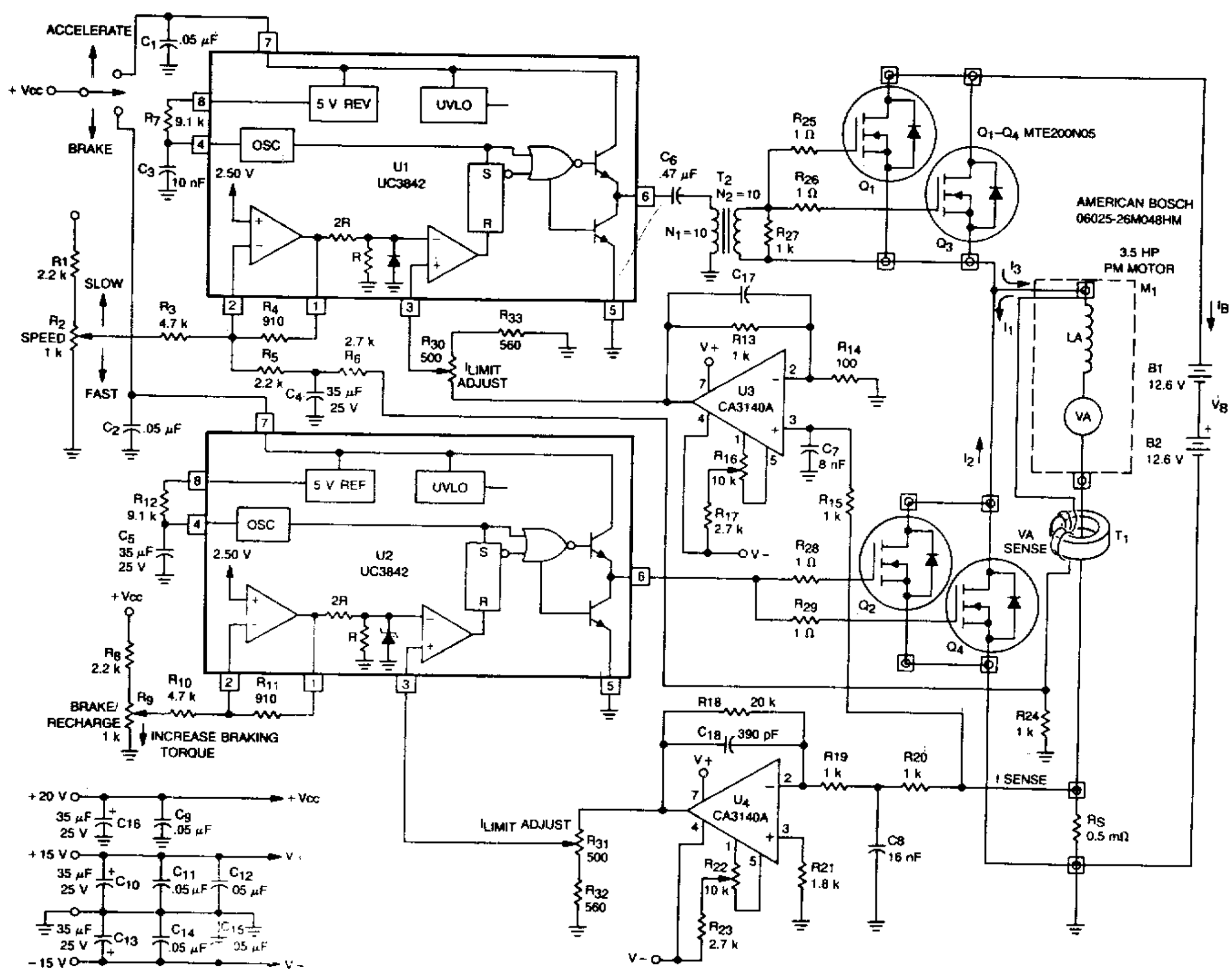


Fig. 62-2

This circuit uses the main drive motor as a generator/brake to recover the battery charge during vehicle braking. When this is done, it can increase the overall range and efficiency of an electric vehicle.

In the accelerate mode, Q1 through Q3 receive gate pulses from U1, an on-line, current-mode, PWM controller IC. Assuming negligible effects of R, when Q1 through Q3 turn on, current I_1 builds up through the motor at a rate of:

$$dI_1 = V_B - V_A$$

where: V_A = Battery voltage
V_B = Motor's back EMF
L_A = Motor inductance in henries

Motor current and torque continue to rise until the voltage on the I_{SENSE} line is greater than I_{RESET}, as determined by the speed potentiometer. At this time, Q1 through Q3 are switched off, current I_2 begins to flow and decreases at a rate of:

$$\frac{dI_2}{dt} = \frac{V_A}{L_A}$$

until the next clock period begins.

As vehicle braking occurs, accelerate PWM IC U1 is switched off and braking PWM IC U2 and Q2 through Q4 are switched on. During this time, the back EMF source voltage causes current I_3 to begin to flow at the rate of:

$$\frac{dI_3}{dt} = \frac{V_A}{L_A}$$

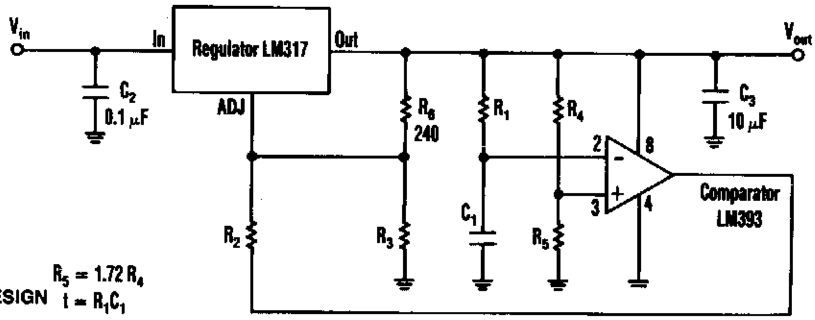
The current I_3 continues to rise until I_{SENSE} is greater than I_{RESET}. Now, Q2 through Q4 are switched off and I_B is forced to flow back into the storage battery, thus energy is recovered.

The braking torque produced by the motor is proportional to the average reverse current that flows through the motor on the duty cycle of Q2 through Q4. The braking force can continue until:

$$V_A = 0$$

For reliable performance, voltage supplies should be independent of the main battery voltage.

START-AND-RUN MOTOR CIRCUIT



ELECTRONIC DESIGN

Fig. 62-3

The timed two-voltage circuit can start and run a small dc motor or solenoid. The input voltage to the LM317 three-terminal regulator ranges from 5 to 40 V, and the output voltage can range from 2 to 36 V. With input voltage V_{IN} initially applied to the input, and capacitor C1 in a discharged state, the LM393 comparator's open-collector output circuit is open-circuited. Then the higher start-up output voltage is:

$$V_{OUT} = 1.25\,[1 + (R_3/240)]$$

At a time t after start-up, when:

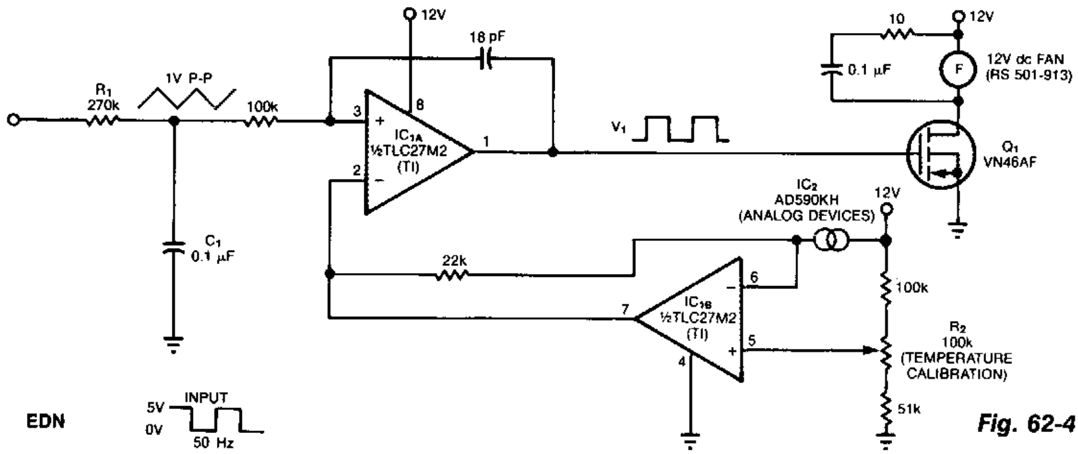
$$t = -R_1C_1\text{Ln}[R_4/(R_4 + R_5)]$$

or:

$$t = -R_1C_1,\ \text{if } R_5 = 1.72R_4$$

the comparator output decreases. At that time, the output voltage switches to a lower value to run the device at its proper operating level.

AUTOMATIC FAN SPEED CONTROLLER



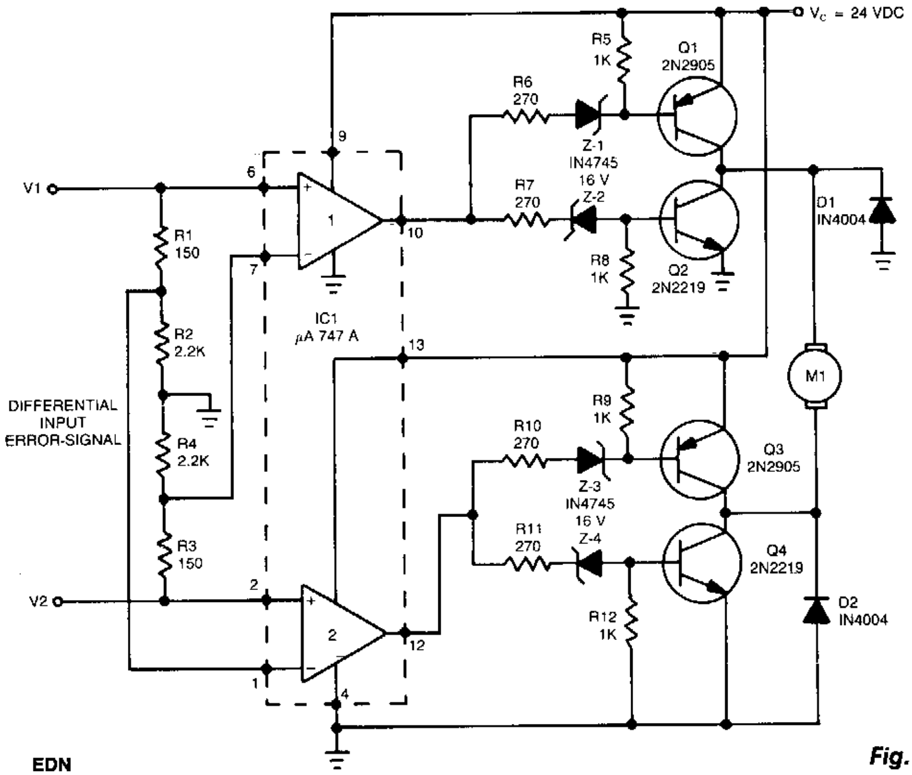
EDN

Fig. 62-4

AUTOMATIC FAN SPEED CONTROLLER (*Cont.*)

The controller circuit can reduce a fan's noise, power consumption, and wear, particularly in the presence of a low, fluctuating ambient temperature. Mount a temperature sensor in the fan's airstream, and the circuit will adjust the fan speed as necessary to maintain a relatively constant sensor temperature. Input components R1 and C1 integrate the input square wave, producing a quasitriangular wave at the noninverting input of op amp IC1A. At this inverting input is a reference voltage that decreases as temperature increases. The two-terminal sensor produces 1 μA/°K. The result is a rectangular wave at the output of IC1A with a duty cycle proportional to absolute temperature. Thus, a rise in temperature triggers a counteracting cooling effect by delivering more power to the fan. To calibrate the system with the sensor at room temperature, simply adjust R2 for a 50% duty cycle at V_1. The fan will switch off at approximately 0°C and will be fully on at 44°C.

EFFICIENT SWITCHING CONTROLLER

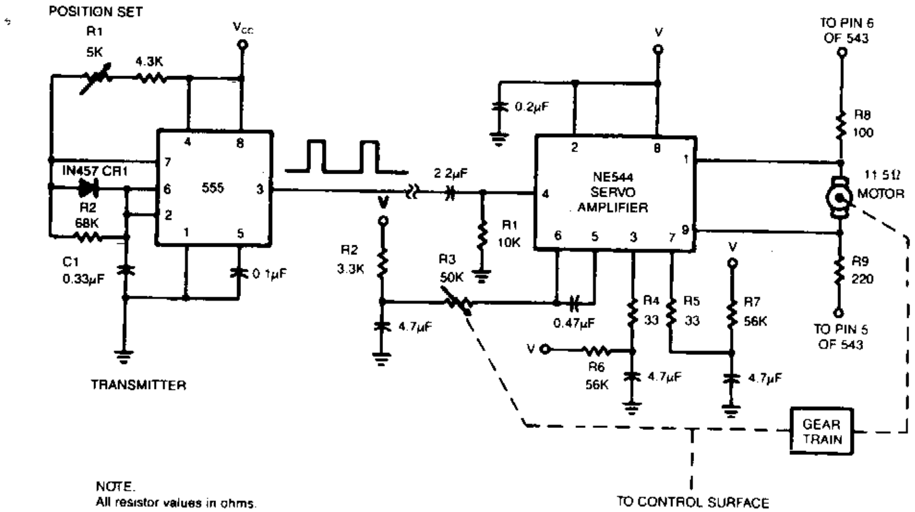


EDN

Fig. 62-5

This high-performance switching controller for a low-power dc servo motor uses a symmetrical complementary-transistor bridge. The bridge acts as a reversing switch between the motor and a single-ended power supply. Since the transistors operate either fully on or completely off, except during a very short transition period, much less heat is dissipated than in linear-amplifier circuits. Damping is provided by the circuit's inherent dynamic braking. Since either maximum or zero voltage is applied to the motor, the dynamic response is faster than that of linear servo drives.

SERVO SYSTEM CONTROLLER

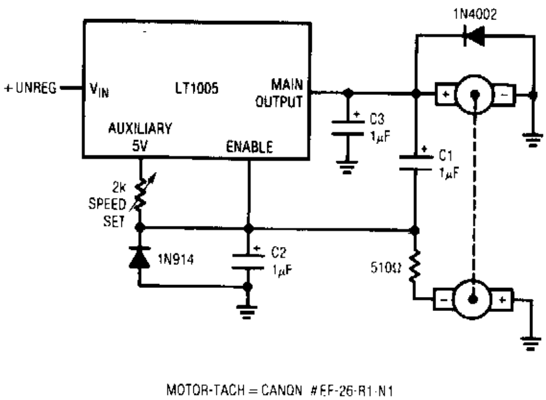


SIGNETICS

Fig. 62-6

To control a servo motor remotely, the 555 needs only six extra components.

SWITCHED-MODE MOTOR-SPEED CONTROLLER



LINEAR TECHNOLOGY CORP.

Fig. 62-7

This circuit uses a tachometer to generate a feedback signal which is compared to a reference supplied by the auxiliary output. When power is applied, the tachometer output is zero and the regulator output comes on, forcing current into the motor. As motor rotation increases, the negative tachometer output pulls the enable pin toward ground. When the enable pin's threshold voltage is reached, the regulator output decreases and the motor slows. C1 provides positive feedback, ensuring clean transitions. In this fashion, the motor's speed is servo-controlled at a point determined by the 2-KΩ potentiometer setting. The regulator free-runs at whatever frequency and duty cycle are required to maintain the enable pin at its threshold. The loop bandwidth and stability are set by C2 and C3. The 1N914 diode prevents the negative output tachometer from pulling the enable pin below ground, and the 1N4002 commutates the motor's negative flyback pulse.

CLOSED-LOOP MOTOR-SPEED CONTROL

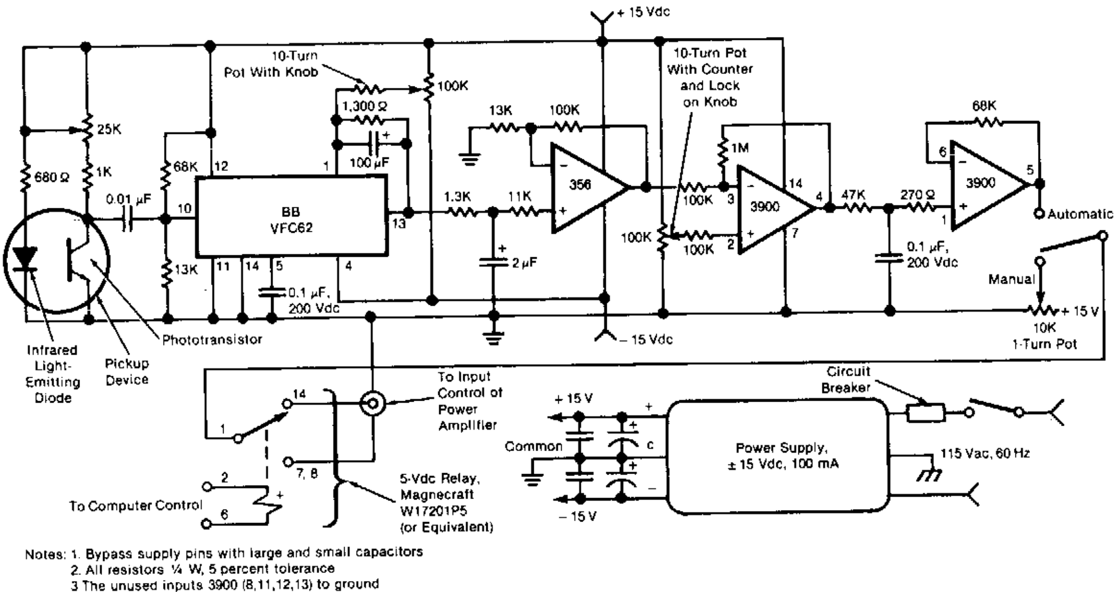


Fig. 62-8

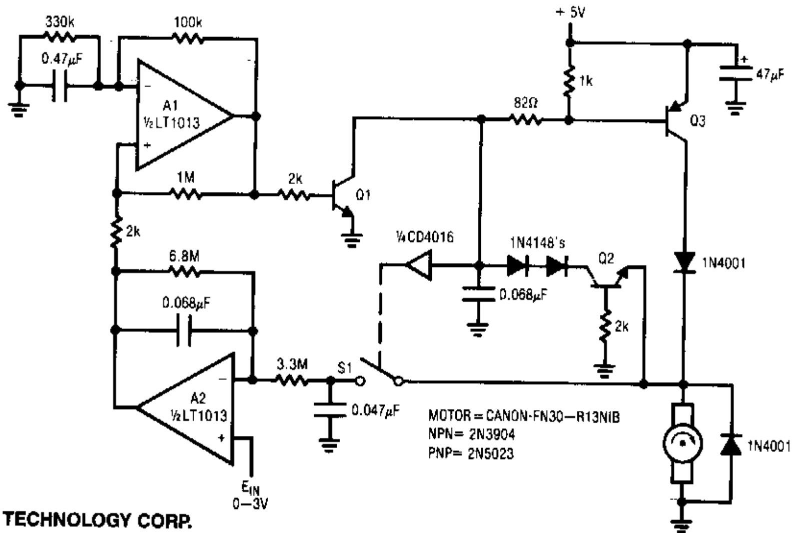
NASA

This electronic motor-speed control circuit is designed to operate in an electrically noisy environment. The circuit includes an optoelectronic pickup device, which is placed inside the motor housing to provide a speed feedback signal. The circuit automatically maintains the speed of the motor at the commanded value.

The pickup device contains an infrared LED and a phototransistor. The radiation from the diode is chopped into pulses by the motor fan blades, which are detected by the phototransistor. The train of pulses from the phototransistor is fed to a frequency-to-voltage converter, the output of which is a voltage proportional to the speed of the motor. This voltage is low-pass filtered, amplified, and compared with a manually-adjustable control voltage that represents the commanded speed.

The difference between the speed-measurement and speed-command signals is amplified and fed as a control voltage to an external power amplifier that drives the motor. A selector switch at the output of the final amplifier of this circuit also enables the operator to bypass the circuit and manually set the control voltage for the external amplifier.

TACHLESS MOTOR-SPEED CONTROLLER

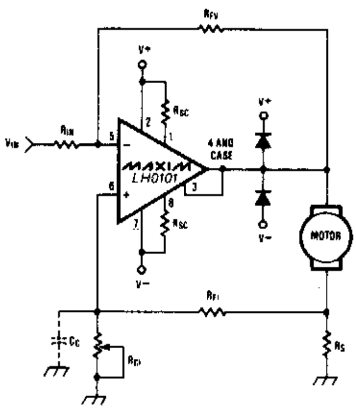


LINEAR TECHNOLOGY CORP.

Fig. 62-9

This circuit is particularly applicable to digitally-controlled systems in robotic and X-Y positioning applications. By functioning from the 5-V logic supply, it eliminates additional motor-drive supplies. The *tachless* feedback saves additional space and cost. The circuit senses the motor's back EMF to determine its speed. The difference between the speed and a set point is used to close a sampled loop around the motor. A1 generates a pulse train. When A1's output is high, Q1 is biased, and Q3 drives the motor's ungrounded terminal. When A1 decreases, Q3 turns off and the motor's back EMF appears after the inductive flyback ceases. During this period, S1's input is turned on, and the 0.047-μF capacitor acquires the back EMF's value. A2 compares this value with the set point and the amplified difference (trace D) changes A1's duty cycle, controlling the motor speed.

CONSTANT-SPEED MOTOR DRIVER



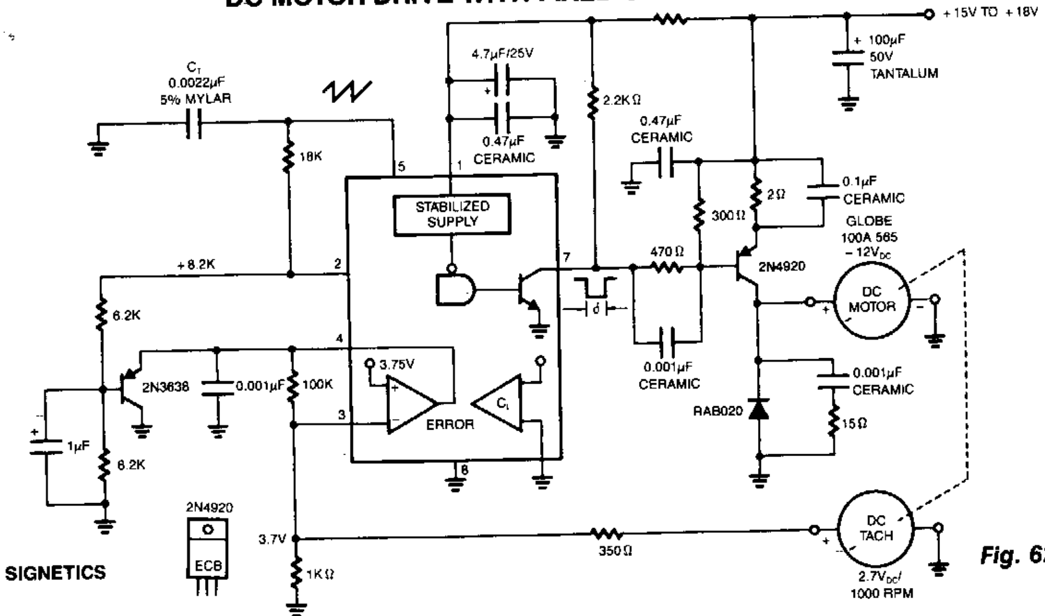
When the torque load on the motor increases, its current increases. This current increase is sensed across R_S, and positive feedback is applied to the noninverting terminal of the LH0101, thereby increasing the motor voltage to compensate for the increased torque load. With the proper amount of positive feedback, the motor-speed variation can be kept below 1% from no load to full load.

MAXIM

ADJUST R_{f1} FOR BEST SPEED vs LOAD REGULATION

Fig. 62-10

DC MOTOR DRIVE WITH FIXED SPEED CONTROL

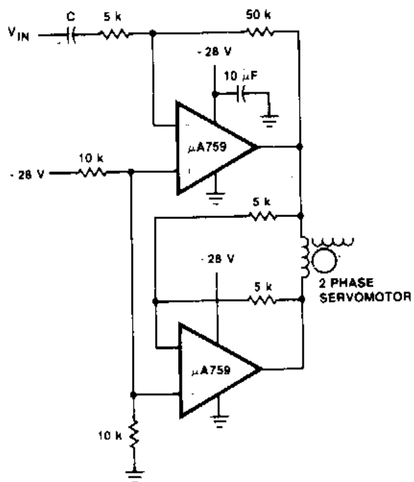


SIGNETICS

Fig. 62-11

The NE5561 provides pulse-proportional drive and speed control based on dc tachometer feedback. This simple switching circuit consists of transistor 2N4920 pnp with a commutation diode used to deliver programmed pulse energy to the motor. A frequency of approximately 20 kHz is used to eliminate audio noise. The dc tach delivers 2.7 V/1000 RPM. Negative feedback occurs when this voltage is applied to the error amplifier of the NE5561. The duty cycle varies directly with load torque demand. The no-load current is ≈ 0.3 A and full load is 0.6 A.

BRIDGE-TYPE AC SERVO AMPLIFIER

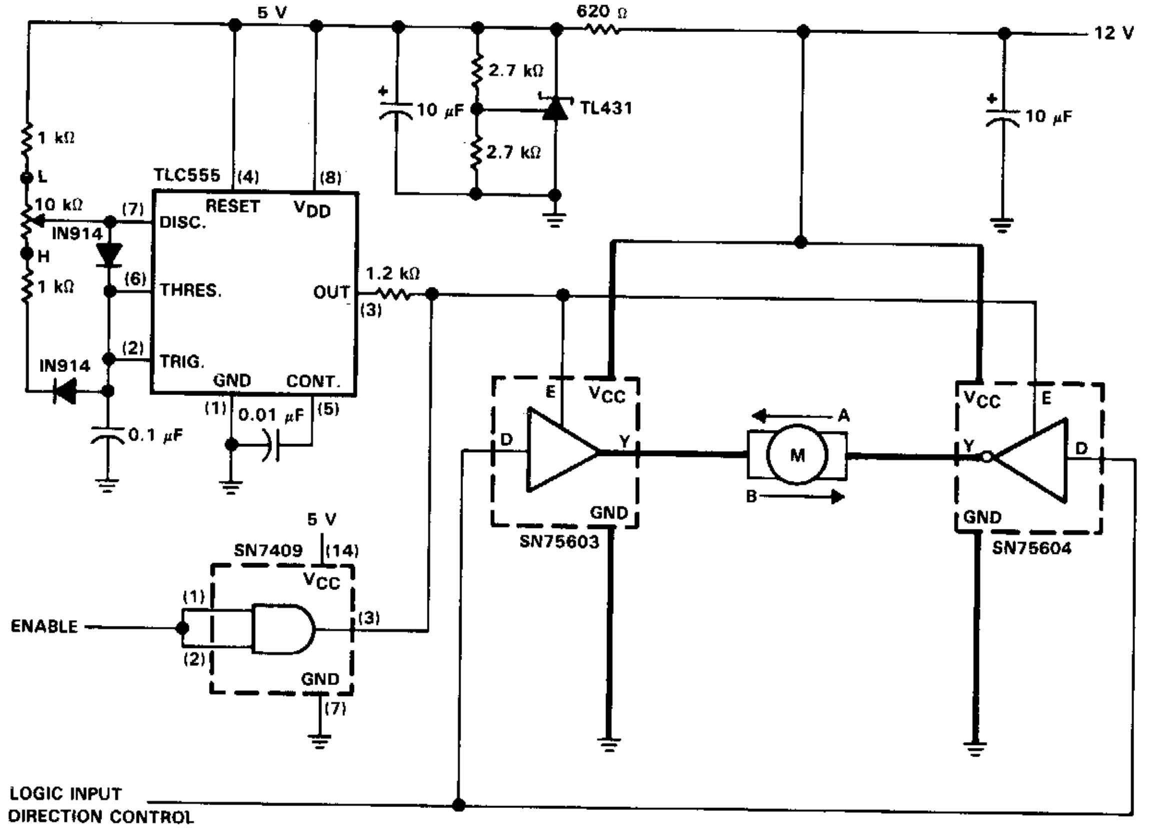


This motor driver circuit uses a μA759 power amplifier to drive a two-phase servomotor.

FAIRCHILD CAMERA AND INSTRUMENT CORP.

Fig. 62-12

SPEED-CONTROLLED REVERSIBLE DC MOTOR DRIVE



Reprinted by permission of Texas Instruments.

Fig. 62-13

The figure illustrates a reversible dc motor drive application with adjustable speed control. The D inputs for these drivers are complementary and can be tied together and driven from the same logic control for bidirectional motor drive. The enables are tied together and driven by a pulse-width-modulated generator providing on duty cycles of 10 to 90% for speed control. A separate enable control is provided through an SN7409 logic gate. See the truth table for this motor controller application.

Definitions for the terms used in the truth table are as follows:

EN	Enable
DC	Direction control
SP.C	Speed control
A	Direction of current—right to left
B	Direction of current—left to right
H	Logic 1 voltage level
L	Logic 0 voltage level
N	Speed control set for narrow pulse width
W	Speed control set for wide pulse width
X	Irrelevant

Truth Table for Motor Control Circuit

EN	DC	SP.C	MOTOR DIRECTION	MOTOR SPEED
L	X	X	OFF	OFF
H	L	N	A	SLOW
H	L	W	A	FAST
H	H	N	B	SLOW
H	H	W	B	FAST

PWM MOTOR CONTROLLER

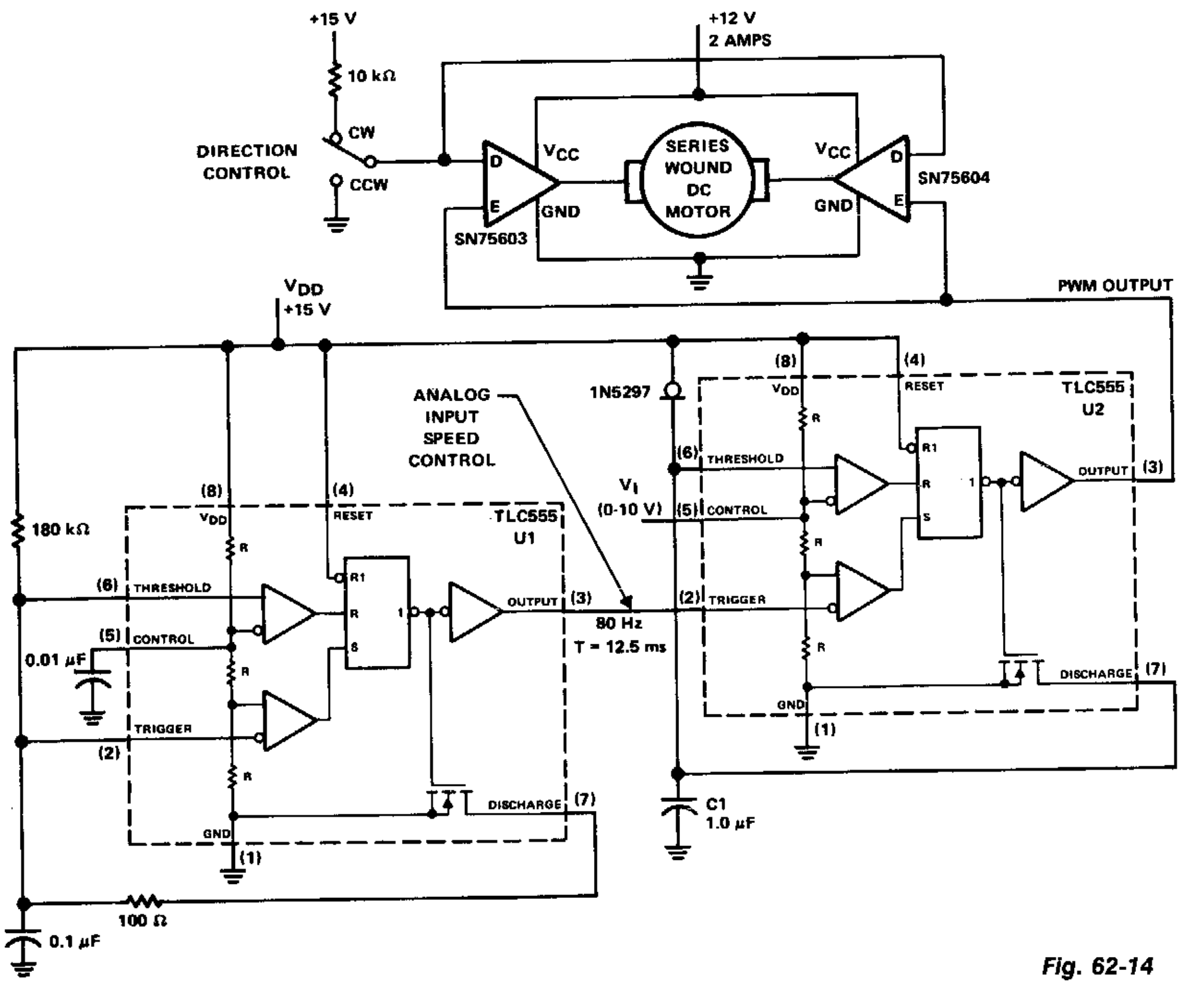


Fig. 62-14

Reprinted by permission of Texas Instruments.

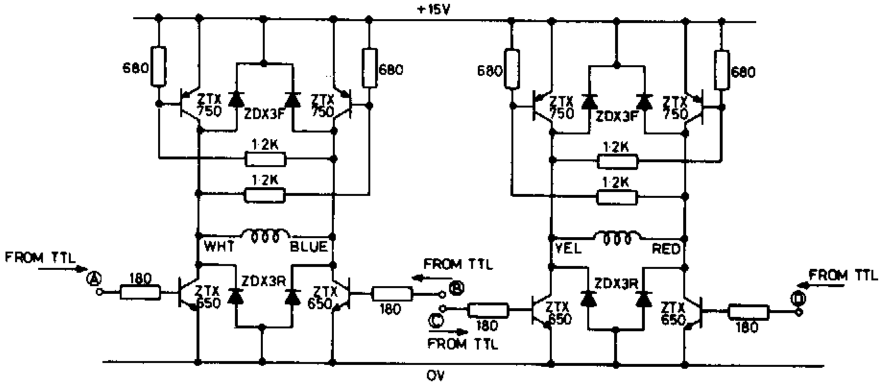
The PWM controller uses complementary half-H peripheral drivers SN75603 and SN75604, with totem-pole outputs rated at 40 V and 2.0 A. These drivers effectively place the motor in a full-bridge configuration, which has the ability to provide bidirectional control.

Timer U1 operates in the astable mode at a frequency of 80 Hz. The 100-Ω discharge resistor results in an 8-µs trigger pulse which is coupled to the trigger input of timer U2. Timer U2 serves as the PWM generator. Capacitor C1 is charged linearly with a constant current of 1 mA from the 1N5297, which is an FET current-regulator diode.

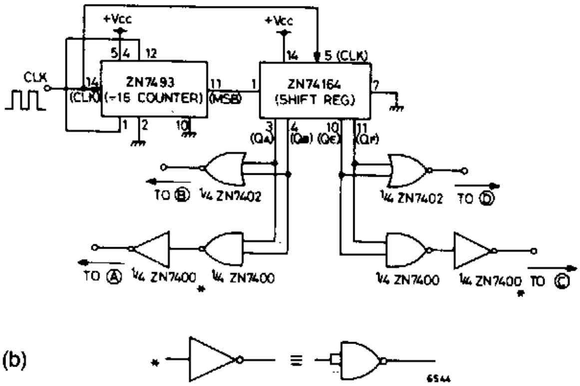
Motor speed is controlled by feeding a dc voltage of 0 to 10 V to control input pin 5 of U2. As the control voltage increases, the width of the output pulse pin 3 also increases. These pulses control the on/off time of the two motor drivers. The trigger pulse width of timer U1 limits the minimum possible duty cycle from U2.

STEPPING MOTOR DRIVE

(a)



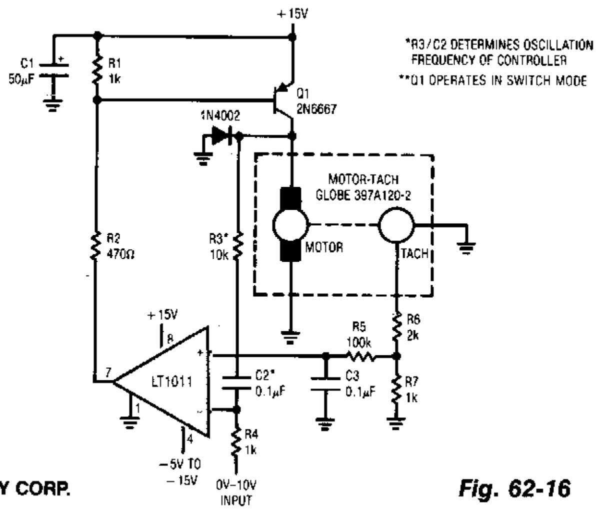
(b)



The circuit shown in Fig. 62-15A is designed to drive a 15-V, two-phase, bipolar stepping motor, providing a bidirectional single level voltage across each winding at currents of up to 9.6 A. The circuit consists of two identical transistor bridge stages employing complementary npn and pnp devices. The transistor conduction sequence is determined by external control logic, and the circuit will interface directly with standard TTL. A suitable control logic system is illustrated in Fig. 62-15B.

ZETEX, formerly FERRANTI ***Fig. 62-15***

HIGH-EFFICIENCY MOTOR-SPEED CONTROLLER



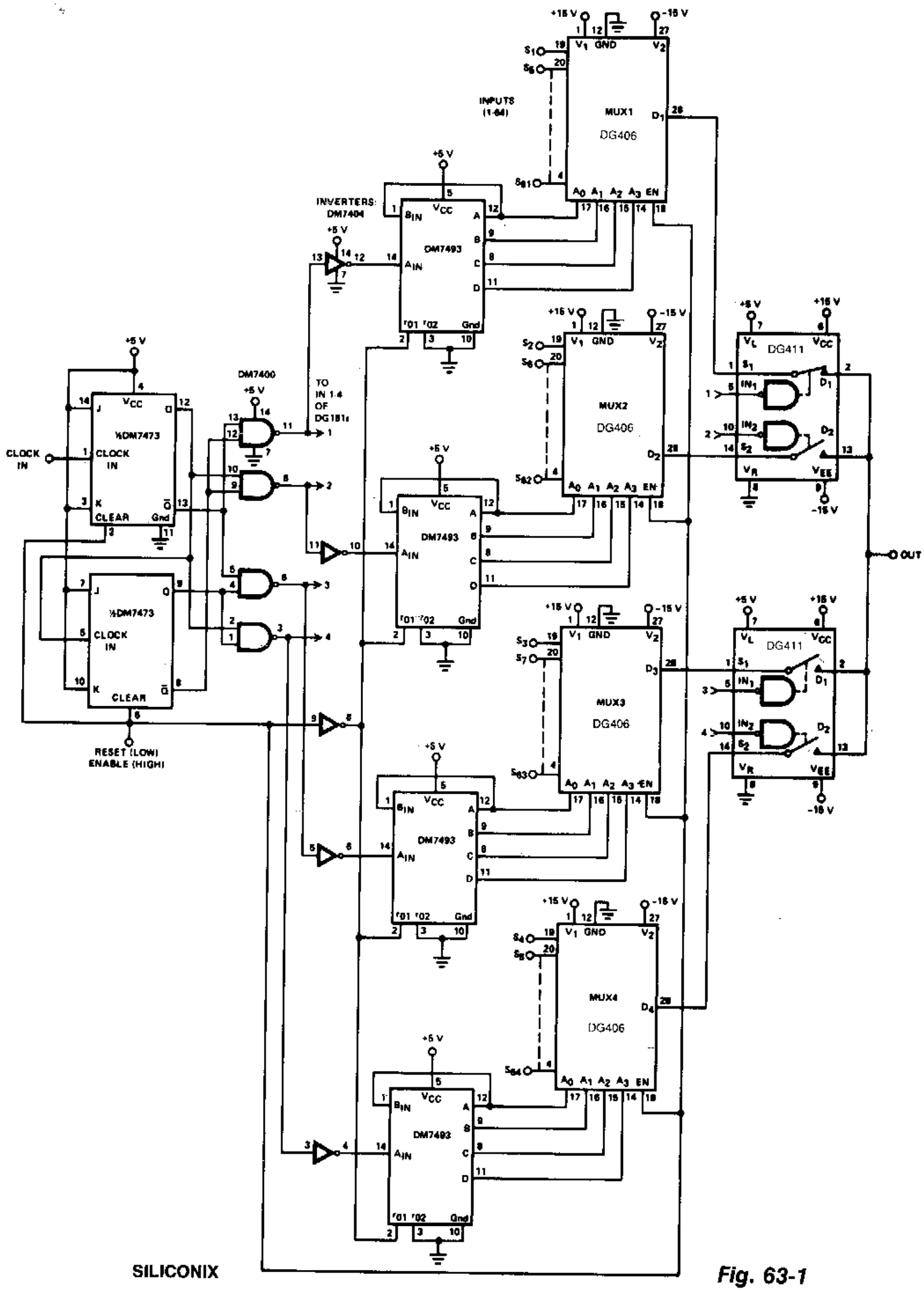
LINEAR TECHNOLOGY CORP. ***Fig. 62-16***

63

Multiplexers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

TWO-LEVEL MULTIPLEXER

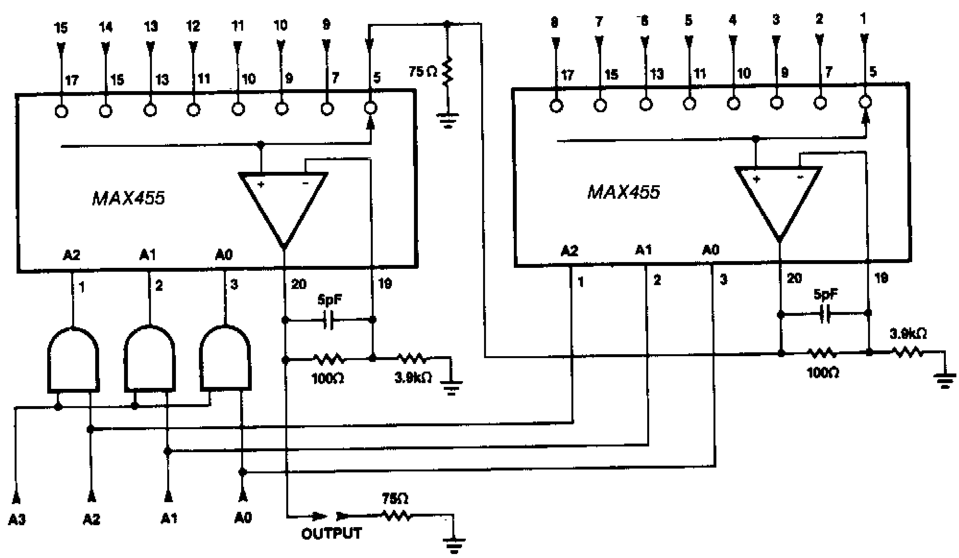


SILICONIX

Fig. 63-1

When a large number of channels are multiplexed, the outputs of two or more multiplexers can be connected together and each multiplexer sequentially enabled. In the inhibit mode, the multiplexer draws less power and its output and inputs act as open circuits. Theoretically, an infinite number of channels can be accommodated in this way; in practice, the accumulated output capacitance and leakage of many paralleled multiplexers limits the speed and accuracy of the system. A much better method is the two-level multiplex system. The two-level system has a bank of high-speed switches at the output which sequentially switch between the four DG406s. Each DG406 is able to switch during the time the other three are being interrogated. The DG406s contribute leakage and capacitance at the output only, when they are switched on by the DG411—1/4 of the time. The two-level multiplex system is very useful in communications links, high-speed interfacing with comparators, or wherever a large number of channels must be multiplexed at high speeds.

1-OF-15 CASCADED VIDEO MUX

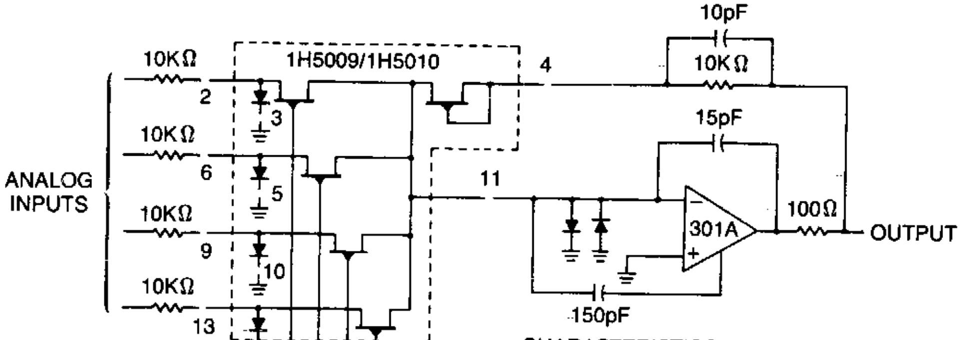


MAXIM

Fig. 63-2

Two MAX455s can be cascaded to form a 1 of 15 video MUX by connecting the output of one MUX to one of the input channels of a second MUX. Although the two devices are usually close to one another, the output of the first MUX should be terminated to preserve its bandwidth.

LOW-COST FOUR-CHANNEL MULTIPLEXER

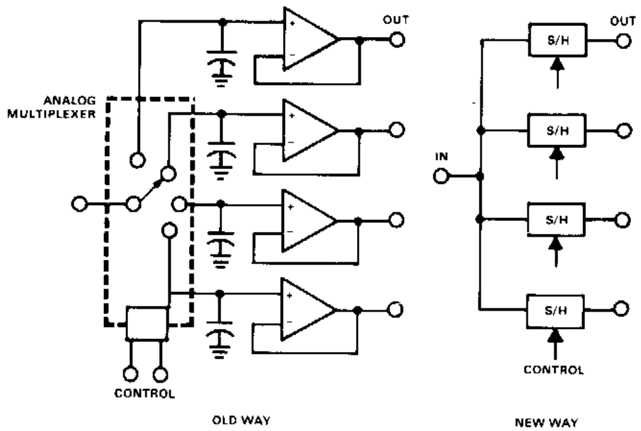


CHARACTERISTICS: SLEW RATE = 10V/μs
GAIN ACCURACY =
.5% (1H5009) TO .05%
(ITS7322). See Table 1.

INTERSIL

Fig. 63-3

DEMULTIPLEXER

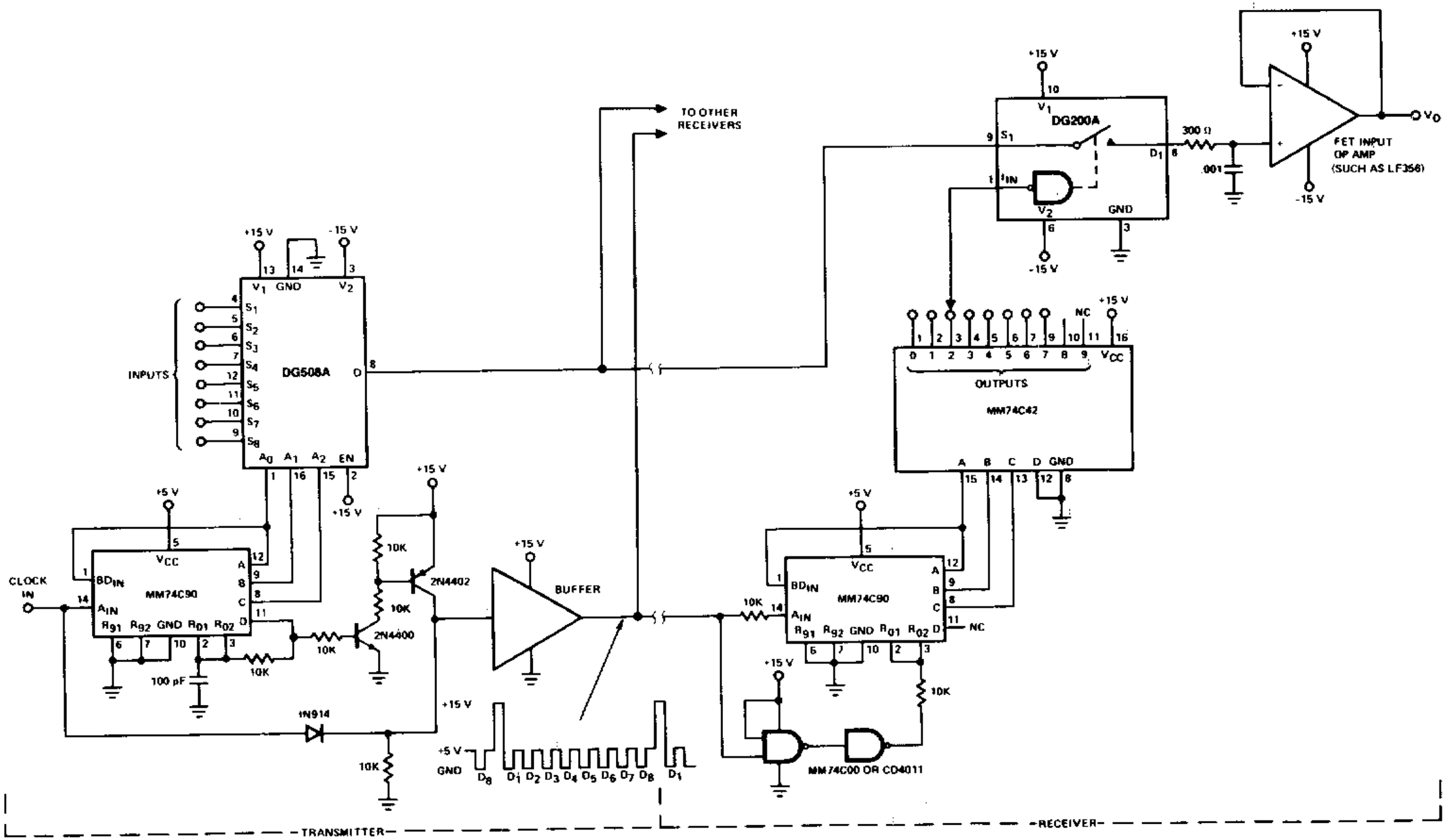


SILICONIX

Fig. 63-4

This circuit reconstructs and separates analog signals which have been time-division multiplexed. The conventional method, shown on the left, has several restrictions, particularly when a short dwell time and a long, accurate hold time is required. The capacitors must charge from a low-impedance source through the resistance and current-limiting characteristics of the multiplexer. When holding, the high-impedance lines are relatively long and subject to noise pickup and leakage. When FET input buffer amplifiers are used for low leakage applications, severe temperature offset errors are often introduced.

1-OF-8 CHANNEL TRANSMISSION SYSTEM

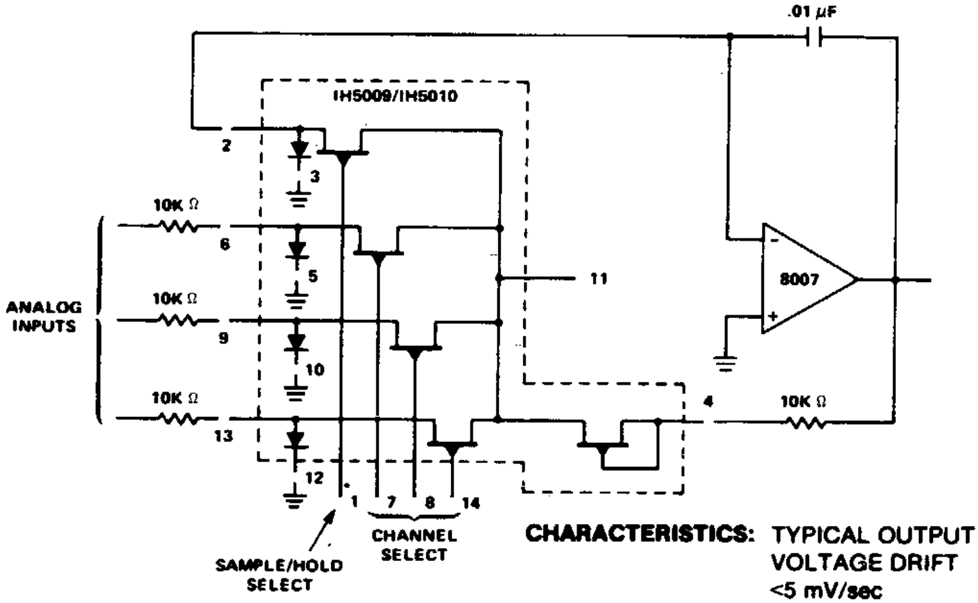


HARRIS

Fig. 63-5

This circuit shows a typical multiplex system intended to carry one of 8 inputs into a remote location. A 5-V pulse train is sent down a separate channel to perform timing and synchronizing functions. A 15-V reset pulse is superimposed on the 5-V clock, which is detected by the MM74C00 in the receiver. Using this system, many remote points can be monitored, one at a time, at any of several locations.

THREE-CHANNEL MULTIPLEXER WITH SAMPLE-AND-HOLD

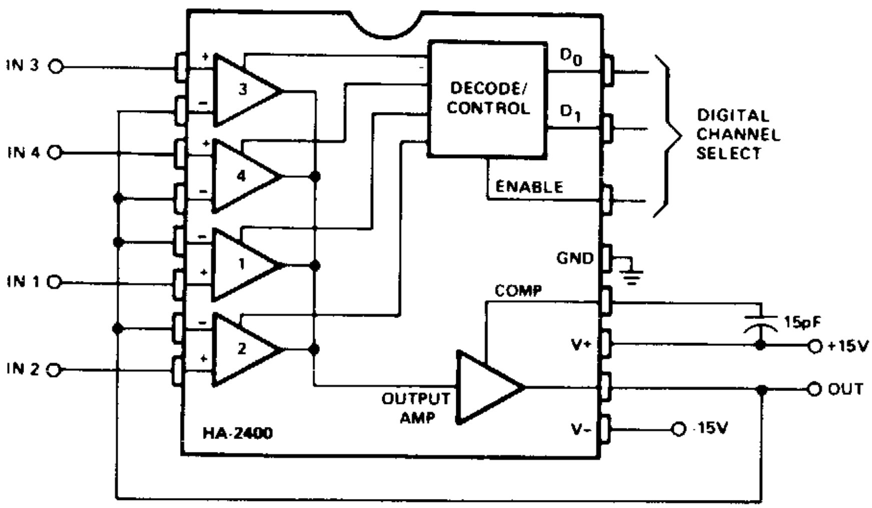


CHARACTERISTICS: TYPICAL OUTPUT VOLTAGE DRIFT <5 mV/sec

INTERSIL

Fig. 63-6

ANALOG MULTIPLEXER WITH BUFFERED INPUT AND OUTPUT



HARRIS

Fig. 63-7

ANALOG MULTIPLEXER WITH BUFFERED INPUT AND OUTPUT (*Cont.*)

This circuit is used for analog signal selection or time division multiplexing. As shown, the feedback signal places the selected amplifier channel in a voltage follower (noninverting unity gain) configuration, and provides very high input impedance and low output impedance. The single package replaces four input buffer amplifiers, four analog switches with decoding, and one output buffer amplifier. For low-level input signals, gain can be added to one or more channels by connecting the (−) inputs to a voltage divider between output and ground. The bandwidth is approximately 8 MHz, and the output will slew from one level to another at about 15.0 V per μs.

Expansion to multiplex 5 to 12 channels can be accomplished by connecting the compensation pins of two or three devices together, and using the output of only one of the devices. The enable input on the unselected devices must be low.

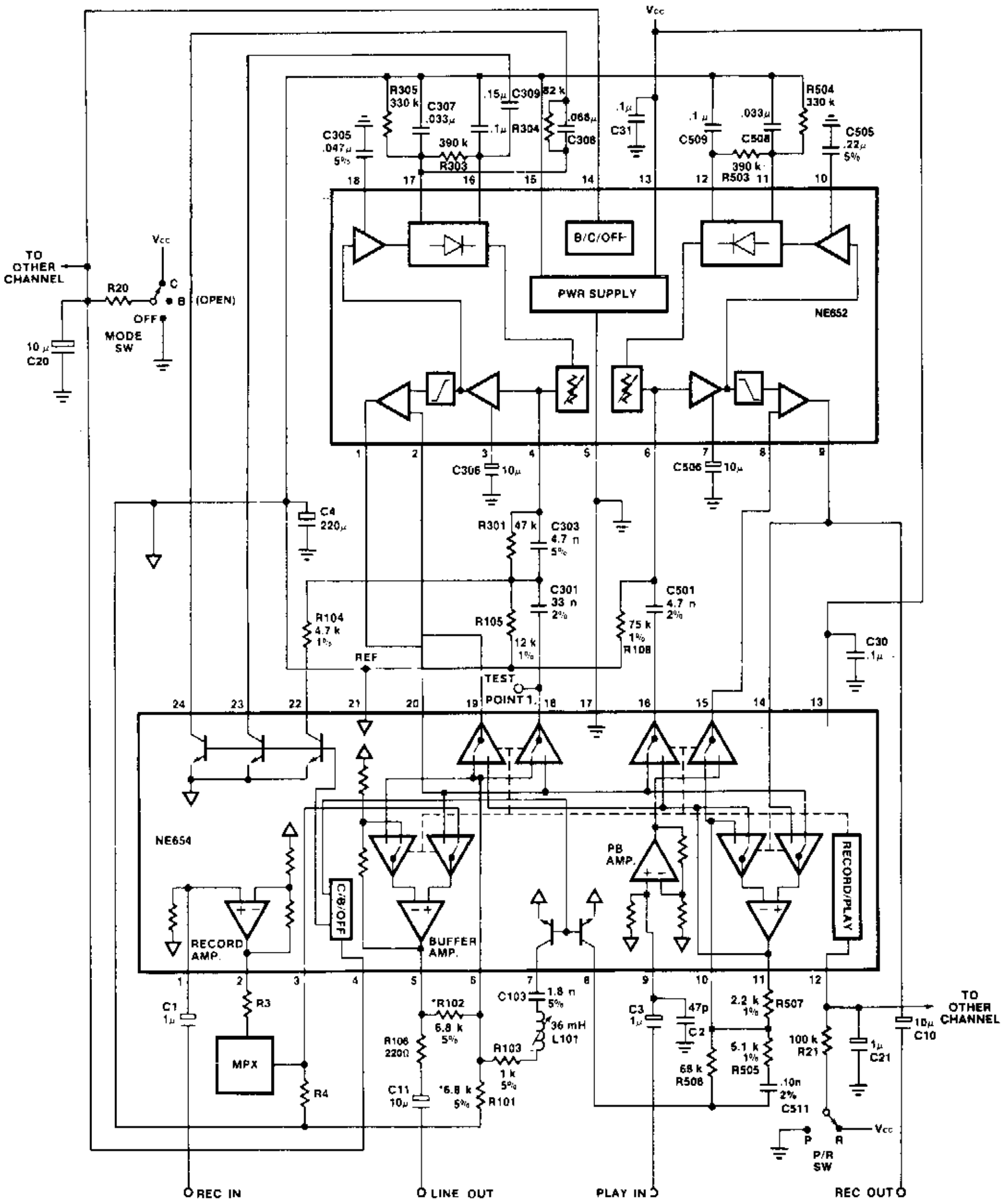
Expansion to 16 or more channels is accomplished easily by connecting outputs of four 4-channel multiplexers to the inputs of another 4-channel multiplexer. Differential signals can be handled by two identical multiplexers addressed in parallel. Inverting amplifier configurations can also be used, but the feedback resistors might cause crosstalk from the output to unselected inputs.

64

Noise Reduction Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

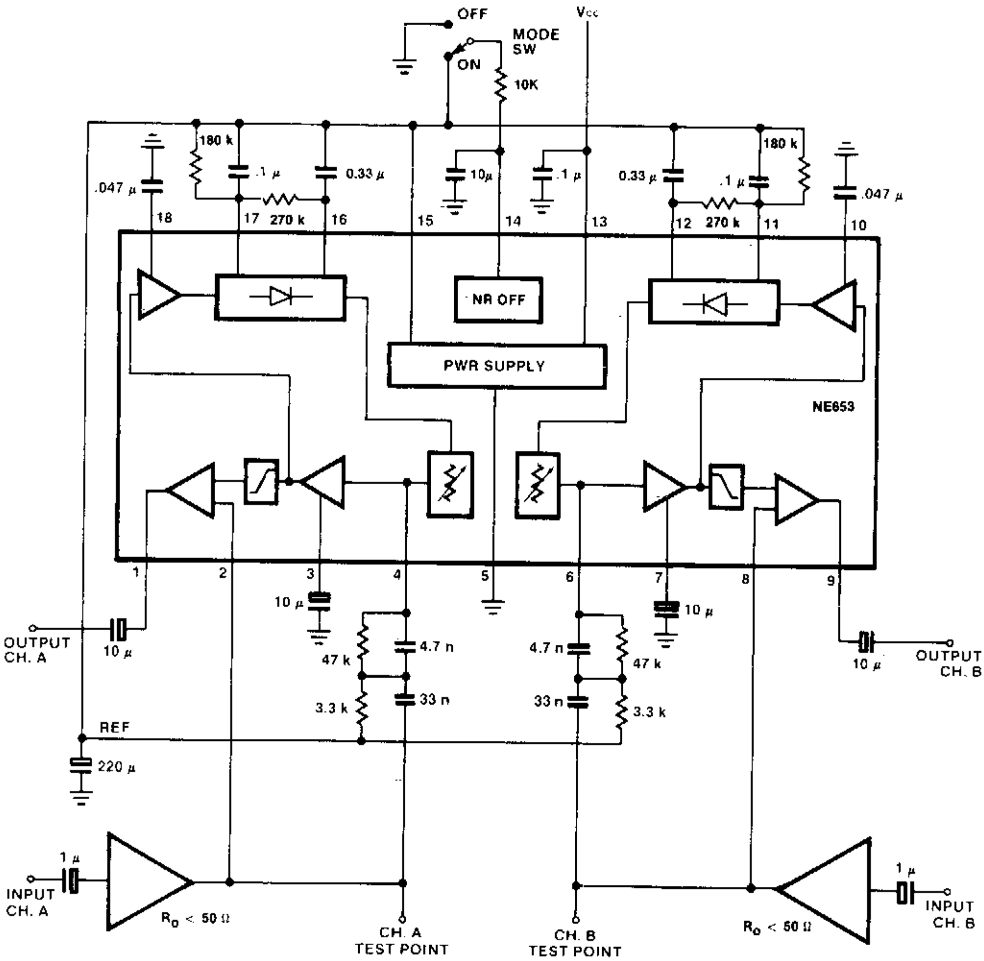
DOLBY B/C NOISE REDUCTION SYSTEM



NOTES:
1. *Line output programming resistors.
2. Split supply operation and coupling capacitors are optional.
3. Time constant for mode switch is optional.
4. Applications info is for reference only. Final design configuration/values are found in relevant Dolby Labs Bulletins and Licensee manuals.
5. R20 value is equal to 6.8KΩ divided by "N" where "N" equals the number of switched channels.
6. R106 is recommended for large capacitive loads on line out.
7. Switches shown in REC position.

SIGNETICS ***Fig. 64-1***

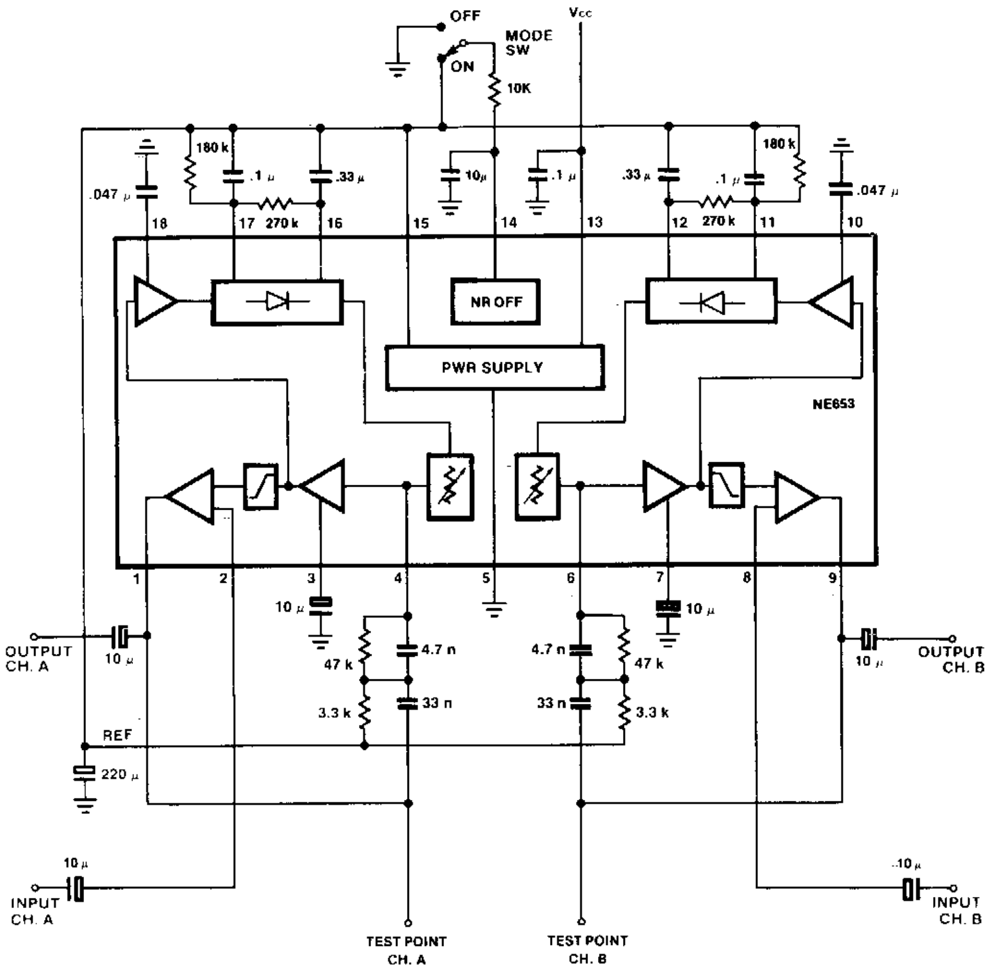
DOLBY B NOISE REDUCTION CIRCUIT IN ENCODE MODE



SIGNETICS

Fig. 64-2

DOLBY B NOISE REDUCTION CIRCUIT IN DECODE MODE



SIGNETICS

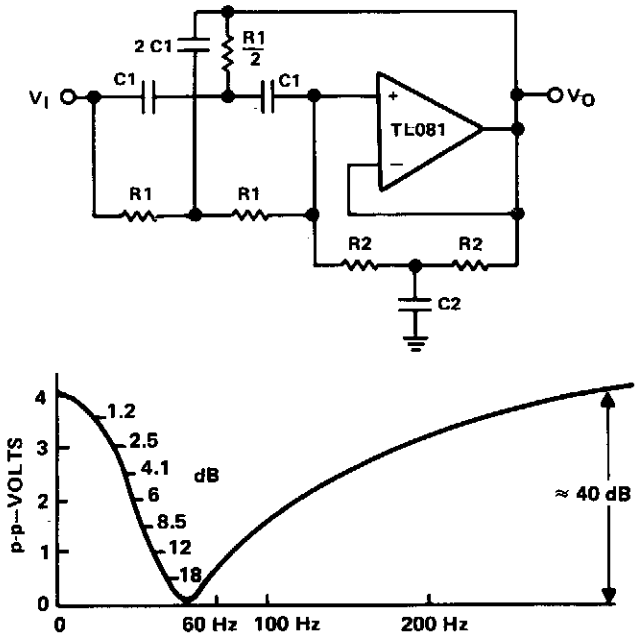
Fig. 64-3

65

Notch Filters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

TWIN-T NOTCH FILTER



60-Hz Twin-T Notch Filter Response

Reprinted by permission of Texas Instruments.

Fig. 65-1

This filter is used to reject or block a frequency or band of frequencies. These filters are often designed into audio and instrumentation systems to eliminate a single frequency, such as 60 Hz. Commercial grade components with 5% – 10% tolerance produce a null depth of at least 30 to 40 dB. When a twin-T network is combined with a TL081 op amp in a circuit, an active filter can be implemented. The added resistor capacitor network, R2 and C2, work effectively in parallel with the original twin-T network, on the input of the filter. These networks set the Q of the filter. The op amp is basically connected as a unity-gain voltage follower. The Q is found from:

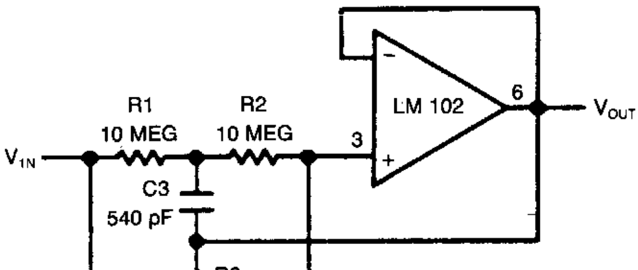
$$Q = \frac{R_2}{2R_1} = \frac{C_1}{C_2}$$

For a 60-Hz notch filter with a Q of 5, it is usually best to pick the C1 capacitor value and calculate the resistor R1. Let C1 = 0.22 μF. Then:

$$R_1 = 12\ \text{K}\Omega$$
$$R_2 = 120\ \text{K}\Omega$$
$$C_2 = 0.047\ \mu\text{F}$$

Standard 5% resistors and 10% capacitors produce a notch depth of about 40 dB, as shown in the frequency response curve.

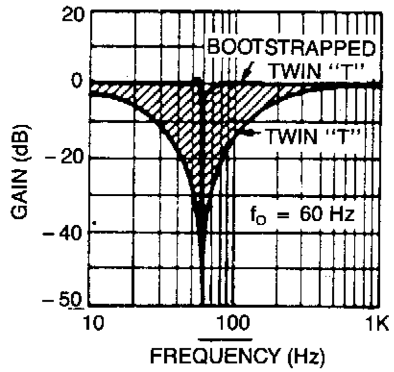
HIGH-*Q* NOTCH FILTER



$$f_O = \frac{1}{2\pi R1 C1} = 60\ \text{Hz}$$

$$R1 = R2 = 2\,R3$$

$$C1 = C2 = \frac{C3}{2}$$



Response of High and Low Q Notch Filter

NATIONAL SEMICONDUCTOR CORP.

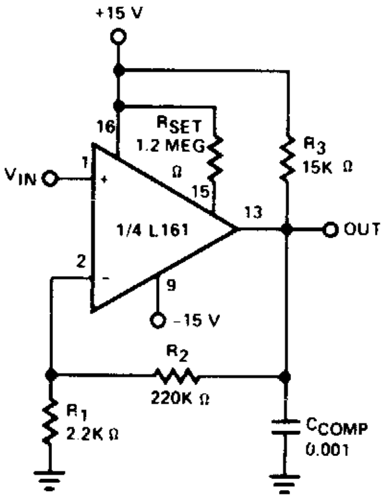
Fig. 65-2

This circuit shows a twin-T network connected to an LM102 to form a high-*Q*, 60-Hz notch filter. The junction of R3 and C3 which is normally connected to ground, is bootstrapped to the output of the follower. Because the output of the follower is a very low impedance, neither the depth nor the frequency of the notch change; however, the *Q* is raised in proportion to the amount of signal fed back to R3 and C3. Shown is the response of a normal twin-T and the response with the follower added.

66

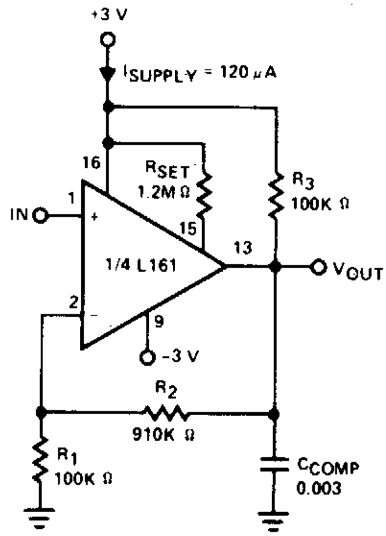
Operational Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.



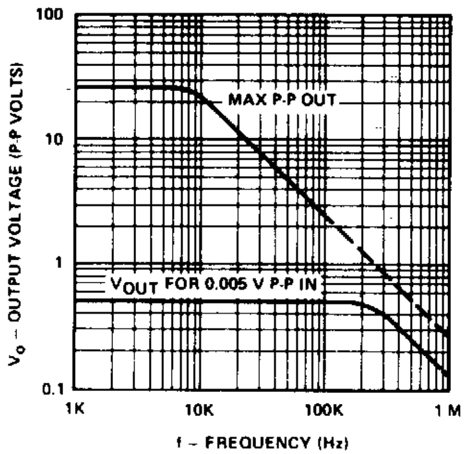
The L161 as a X100 Operational Amplifier

(A)



A Micropower X10 Op Amp

(B)



Frequency Response and Maximum Output for the X100 Op Amp

SILICONIX

Fig. 66-1

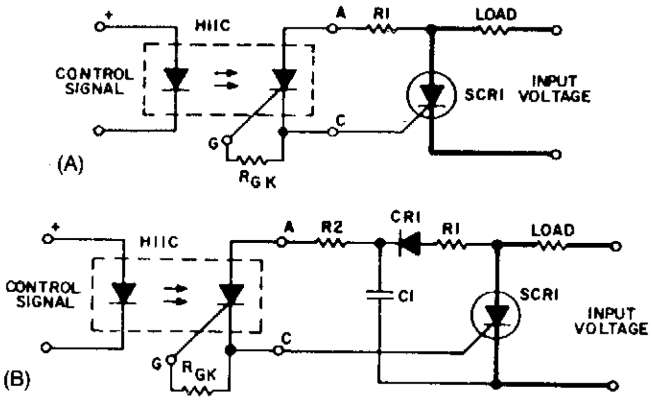
This is a single gain-of-100 amplifier with a gain-bandwidth product of 20 MHz! The primary limitation in the performance is the low slew rate (0.3 V/μs) imposed by I_{OH} charging C_{COMP}. The effects of slew rate and compensation are shown. A lower gain amplifier requires a larger C_{COMP}, which in turn further reduces slew rate. For this reason, it might actually be advantageous in certain areas to lower the gain by placing a resistive divider at the input rather than raising R_I. Figure 66-1B shows a 700-μW, X10 op amp whose slew rate is 0.02 V/μs and is 3 dB down at 100 kHz.

67

Optically-Coupled Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

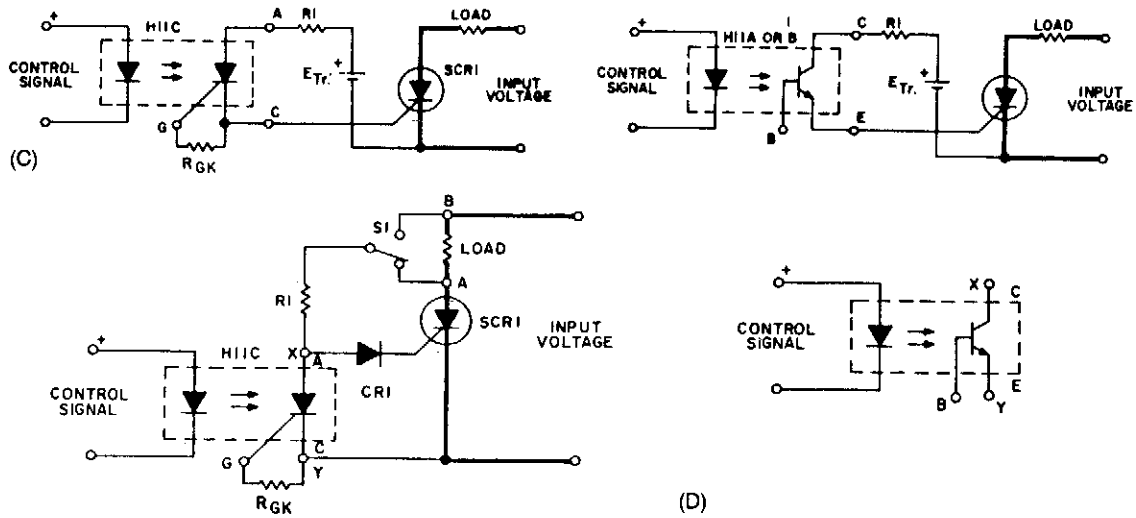
HIGH-VOLTAGE AC SWITCHER



A basic circuit to trigger an SCR is shown in Fig. 67-1A. This circuit has the disadvantage that the blocking voltage of the photon-coupler output device determines the circuit-blocking voltage, irrespective of higher main SCR capability.

Adding capacitor C1 to the circuit, as shown in Fig. 67-1B, will reduce the *dV/dt* seen by the photon-coupler output device. The energy stored in C1, when discharged into the gate of SCR1, will improve the *di/dt* capability of the main SCR.

Using a separate power supply for the coupler adds flexibility to the trigger circuit; it removes the limitation of the blocking voltage capability of the photon-coupler output device. The flexibility adds cost and more than one power supply might be necessary for multiple SCRs if no common reference points are available.



Photon Coupler With SCR – Output

Photon Coupler With Transistor Output (connect in place of SCR coupler)

In Fig. 67-1C, R1 can be connected to Point A, which will remove the voltage from the coupler after SCR1 is triggered, or to Point B so that the coupler output will always be biased by input voltage. The former is preferred since it decreases the power dissipation in R1. A more practical form of SCR triggering is shown in Fig. 67-1F. Trigger energy is obtained from the anode supply and stored in C1. Coupler voltage is limited by the zener voltage. This approach permits switching of higher voltages than the blocking voltage capability of the output device of the photon coupler. To reduce the power losses in R1 and to obtain shorter time constants for charging C1, the zener diode is used instead of a resistor.

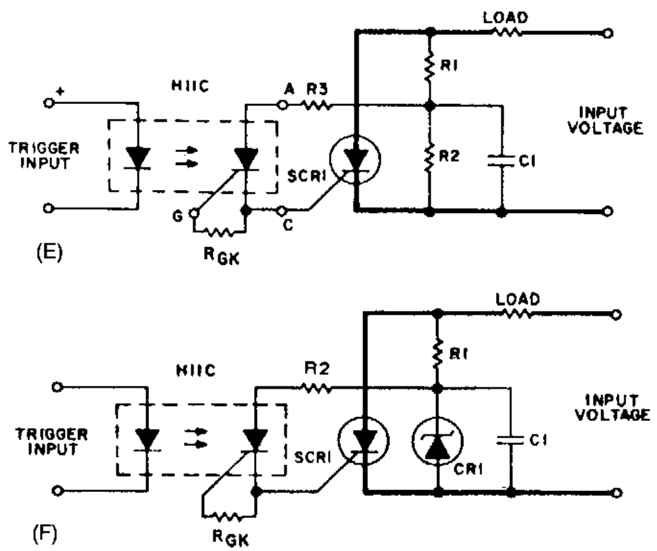


Fig. 67-1

A guide for selecting the component values would consist of the following steps:

- Choose C1 in a range of 0.05 to 1 μF. The maximum value might be limited by the recharging time constant $(R_L + R_1)\ C_1$ while the minimum value will be set by the minimum pulse width required to ensure SCR latching.
- R2 is determined from peak gate current limits, if applicable, and minimum pulse width requirements.
- Select a zener diode. A 25-V zener is a practical value, since this will meet the usual gate requirement of 20 V and 20 Ω. This diode will also eliminate spurious triggering because of voltage transients.
- Photon coupler triggering is ideal for the SCR's driving inductive loads. By ensuring that the LASCR latches on, it can supply gate current to SCR1 until it stays on.
- Component values for dc voltage are easily computed from the following formulae:

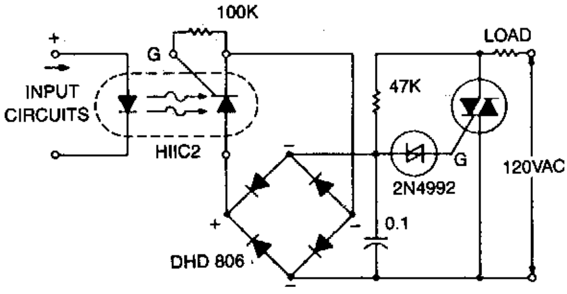
$$R_1 = \frac{E_{IN} - V_Z}{I_G}$$

where: V = zener voltage

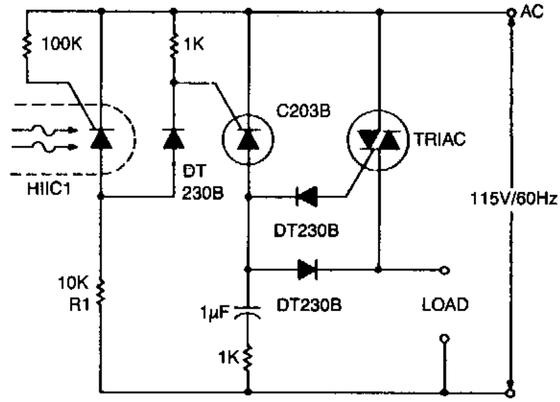
$P_{(R1)} = I_G \cdot (E_{IN} - V_Z)$

$P_{(ZENER)} = I_G \cdot V_Z$

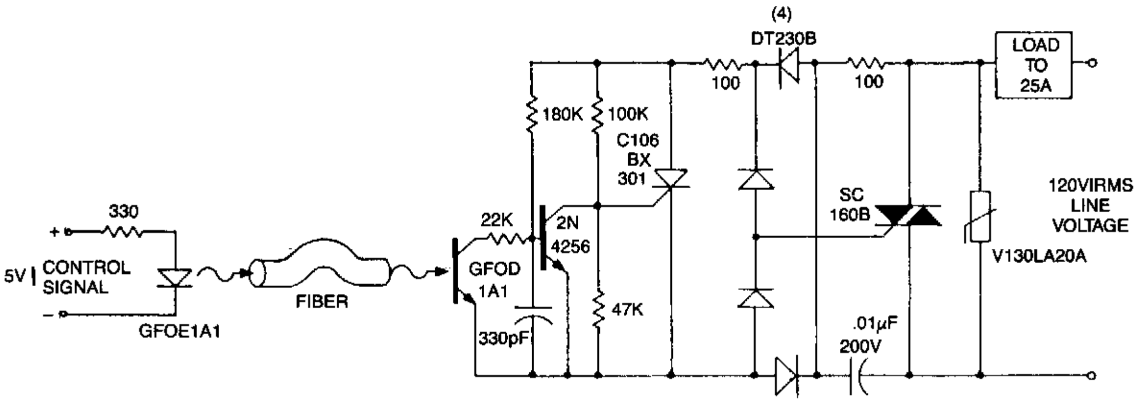
USING A SCR COUPLER



NORMALLY CLOSED
CONTACT ZVS RELAY CIRCUIT



NORMALLY CLOSED INTEGRAL CYCLE,
ZERO VOLTAGE SWITCHING, CONTACT CIRCUIT



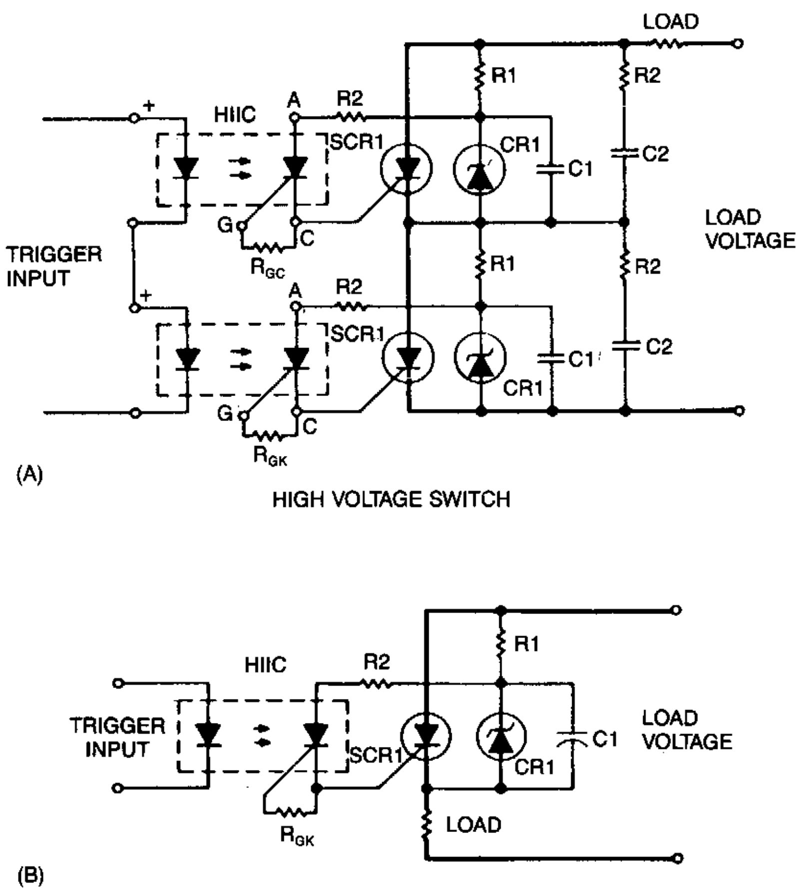
GE

25A FIBER OPTIC ZVS AC SOLID STATE RELAY

Fig. 67-2

This circuit is effective for lamp and heater loads. Some circuits driving reactive loads require integral cycling and zero-voltage switching—when an identical number of positive and negative half cycles of voltage are applied to the load during a power period. The circuit, although not strictly a relay because of the three-terminal power connection, performs the integral cycle ZVS function when interfaced with the previous coil circuits.

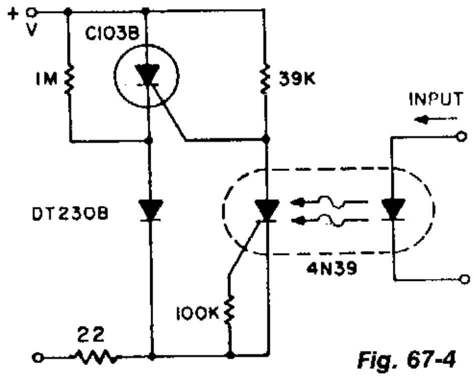
Fiber optics offers advantages in power control systems. Electrical signals do not flow along the nonconducting fiber, minimizing shock hazard to both operator and equipment. EMI/RFI pick up on the fiber is nonexistent—although high gain receiver circuits might require shielding, eliminating noise pick-up errors caused by sources along the cable route. Both ac and dc power systems can be controlled by fiber optics using techniques similar to the optoisolator solid-state relay. Triac triggering is accomplished through the C106BX301, a low gate trigger current SCR, switching line voltage derived current to the triac gate via the full-wave rectifier bridge. The primary difference between fiber optics solid-state relay circuits and optoisolator circuits is the gain; photo currents are much smaller.



GE *Fig. 67-3*

Snubber circuit R2C2, as shown, might be necessary since R1 and C1 are tailored for optimized triggering and not for *dV/dt* protection. Fiber-optic pairs can be used with discrete SCRs to switch thousands of volts. A photon coupler with a transistor output will limit the trigger-pulse amplitude and rise time because of CTR and saturation effects. Using the H11C1, the rise time of the input pulse to the photon coupler is not critical, and its amplitude is limited only by the H11C1 turn-on sensitivity. The load can also be connected to the cathode as illustrated in Fig. 67-3B.

NORMALLY CLOSED HALF-WAVE ZVS CONTACT CIRCUIT

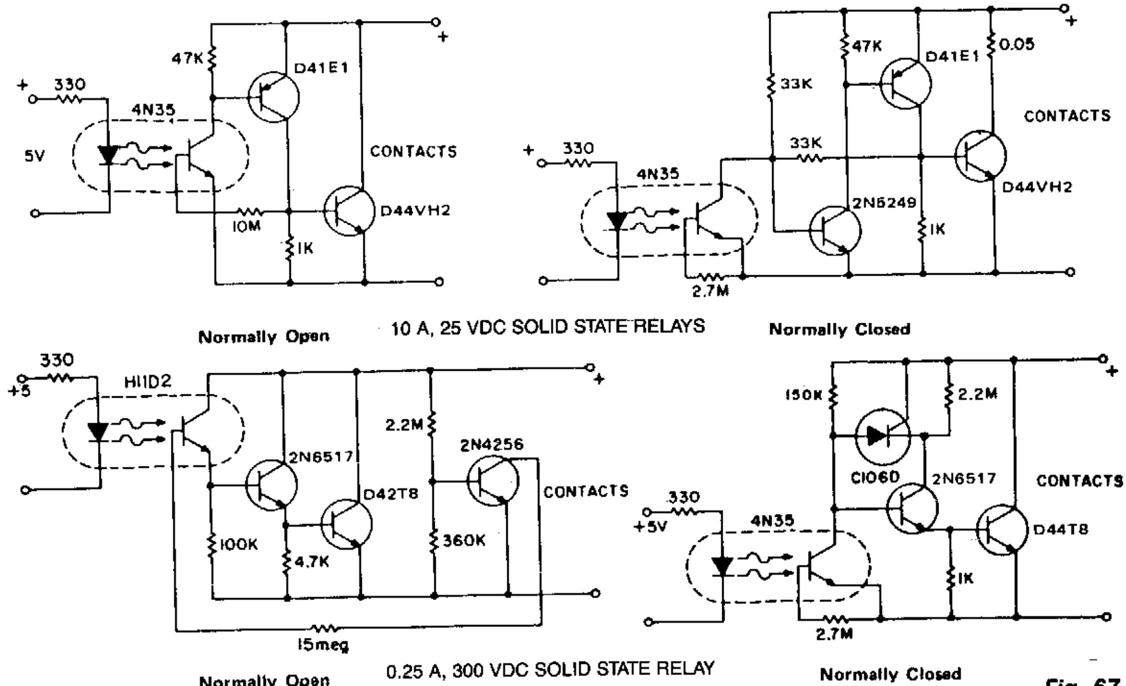


GE

Fig. 67-4

A normally closed contact circuit that provides zero-voltage switching is designed around the 4N39 SCR optocoupler. The circuit illustrates the method of modifying the normally open contact circuit by using the photo SCR to hold off the trigger SCR.

NORMALLY OPEN AND NORMALLY CLOSED DC SOLID-STATE RELAYS

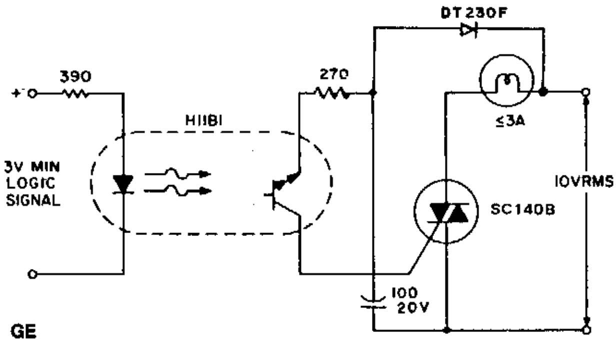


GE

Fig. 67-5

The phototransistor and photodarlington couplers act as dc relays in saturated switching at currents up to 5 mA and 50 mA, respectively. When higher currents or higher voltage capabilities are required, additional devices are required to buffer or amplify the photocoupler output. The addition of hysteresis to provide fast switching and stable pick-up and drop-out points can be easily implemented simultaneously. These circuits provide several approaches to implement the dc relay function and serve as practical, cost-effective examples.

INDICATOR LAMP DRIVER

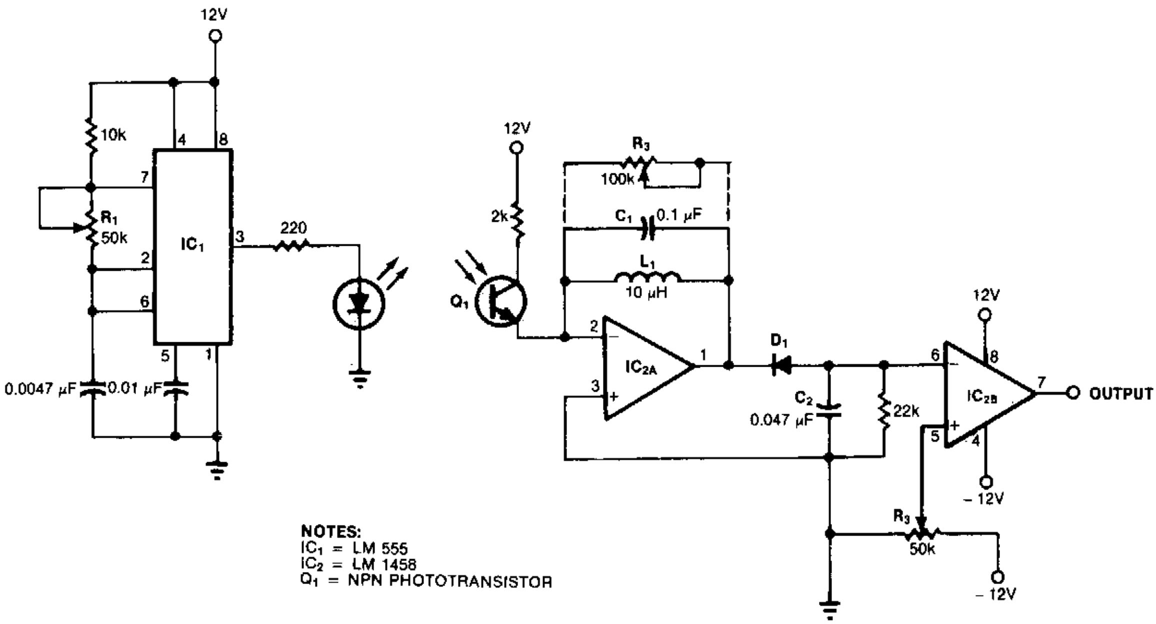


GE

Fig. 67-6

A simple solid-state relay circuit drives the 10-Vac telephone indicator lamps from logic circuitry, while maintaining complete isolation between the 10-V line and the logic circuit.

AMBIENT-LIGHT-IGNORING OPTICAL SENSOR

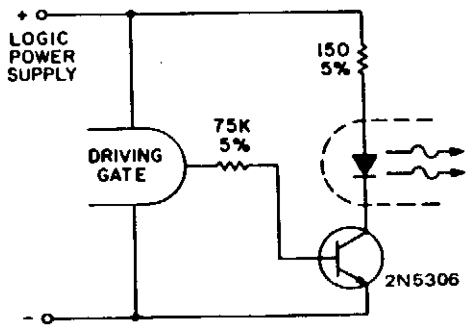


EDN

Fig. 67-7

A resonance-tuned narrow-band amplifier reduces this optical object detector's sensitivity to stray light. C1 and L1 in IC2A's feedback loop cause the op amp to pass only those frequencies at or near the LED's 5-kHz modulation rate. IC2B's output increases when the received signal is sufficient to drop the negative voltage across C2 below the reference set by R2.

OPTICAL CMOS COUPLER

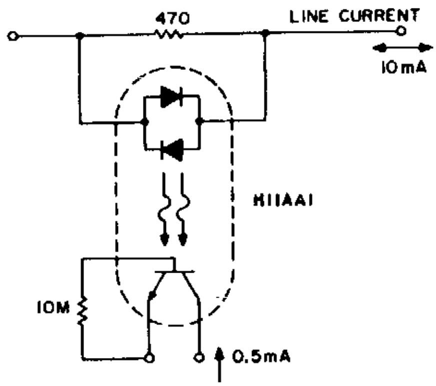


GE

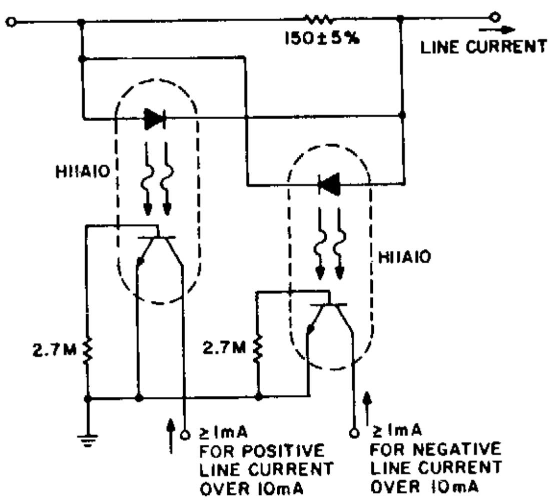
Fig. 67-8

Since standard CMOS logic operates down to 3-V supply voltages and is specified as low as 30 μA maximum current sinking/sourcing capability, it is necessary to use a buffer transistor to provide the required current to the IRED if CMOS is to drive the optocoupler. As in the case of the low output TTL families, the H74A output can drive a multiplicity of CMOS gate inputs or a standard TTL input given the proper bias of the IRED. A one-logic stage drives the IRED on. This circuit will provide worst-case drive criteria to the IRED for logic supply voltages from 3 to 10 V, although lower power dissipation can be obtained by using higher value resistors for high supply voltages. If this is desired, the worst-case drive must be supplied to the IRED with minimum supply voltage, minimum temperature and maximum resistor tolerances, gate saturation resistance, and transistor saturation voltages applied. For the H74 devices, minimum IRED current at worst-case conditions, zero logic state output of the driving gate, is 6.5 mA and the H11L1 is 1.6 mA.

LINE-CURRENT DETECTOR



POLARITY INSENSITIVE LINE CURRENT DETECTOR



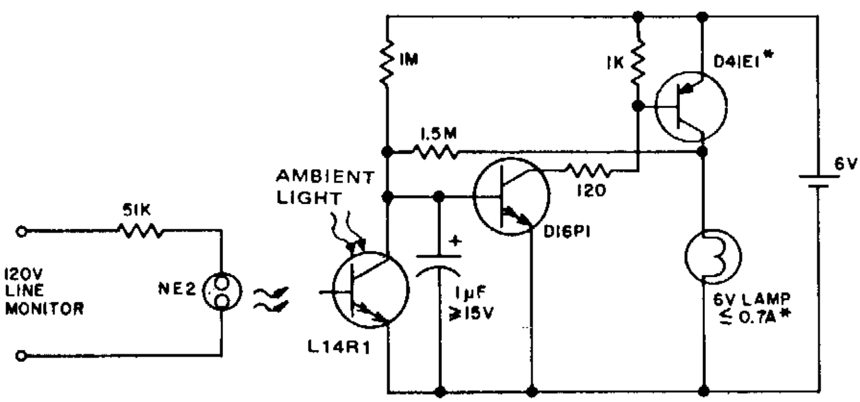
POLARITY INDICATING LINE CURRENT DETECTOR

GE

Fig. 67-9

Detection of line-current flow and indicating the flow to an electrically remote point is required in line status monitoring at a variety of points in the telephone system and auxiliary systems. The line should be minimally unbalanced or loaded by the monitor circuit, and relatively high levels of 60-Hz induced voltages must be ignored. The H11AA1 allows line currents of either polarity to be sensed without discrimination and will ignore noise up to approximately 2.5 mA. In applications where greater noise immunity or polarity-sensitive line-current detection is required, the H11A10 threshold coupler can be used. This phototransistor coupler is specified to provide a minimum 10% current transfer ratio at a defined input current, while leaking less than 50 μA at half that input current over the full $-55°C$ to $+100°C$ temperature range. The input current range, at which the coupler is on, is programmable by a single resistor from 5 to 10 mA.

LINE-OPERATED POWER OUTAGE LIGHT

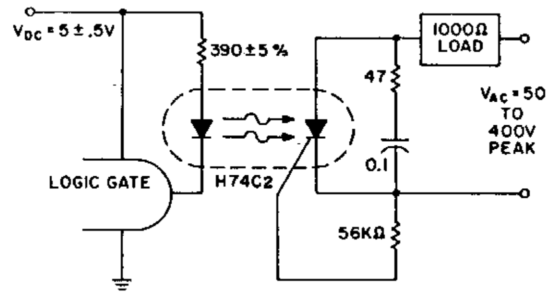


GE

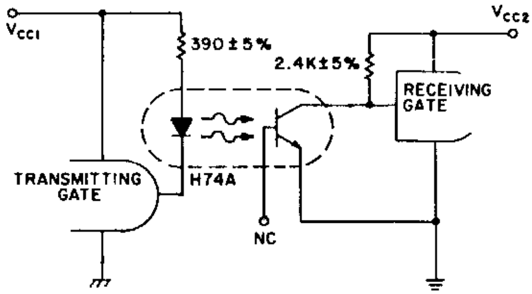
Fig. 67-10

This circuit provides emergency lighting during a power outage. The phototransistor should be positioned to maximize coupling of both neon light and ambient light into the pellet, without allowing self-illumination from the 6-V lamp. Many circuits of this type also use line voltage to charge the battery.

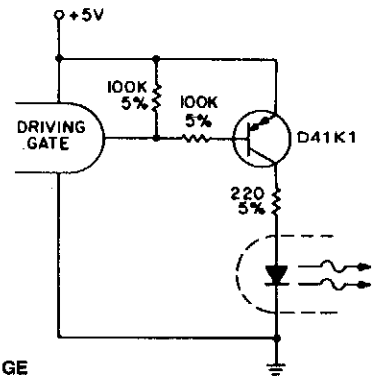
OPTICAL TTL COUPLER



LOGIC TO POWER COUPLING H74 BIAS CIRCUIT



LOGIC TO LOGIC COUPLERS H74A1 BIAS CIRCUIT



GE

IRED DRIVE FROM LOW POWER, MSI AND LSI TTL

Fig. 67-11

For higher speed applications, up to 1-MHz NRZ, the Schmitt-trigger output H11L series optoisolator provides many features. The 1.6-mA drive current allows fan-in circuitry to drive the IRED, while the 5-V, 270-Ω sink capability and 100-ns transition times of the output add to the logic coupling flexibility.

ZERO-VOLTAGE SWITCHING, SOLID-STATE RELAY WITH ANTIPARALLEL SCR OUTPUT

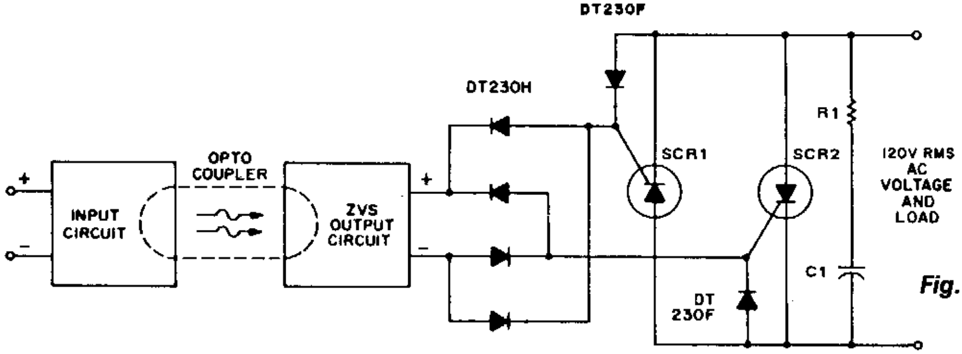
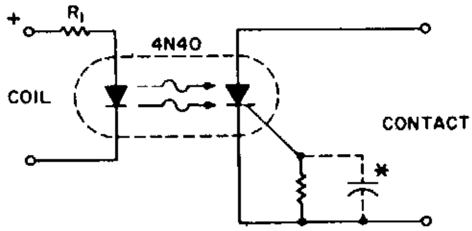


Fig. 67-12

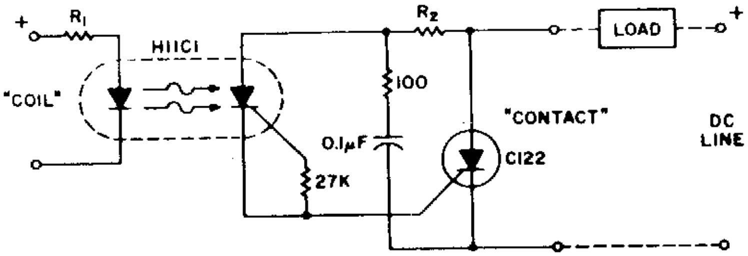
GE

A higher line voltage can be used if the diode, varistor, ZVS, and power thyristor settings are at compatible levels. For applications beyond triac current ratings, antiparallel SCRs might be triggered by the ZVS network.

DC LATCHING RELAY



DC LATCHING RELAY CIRCUIT



COIL VOLTAGE	6	12	24	48	120	V
RI VALUE	470	1.1K	2.4K	4.7K	12K	Ω

LINE VOLTAGE	12	24	48	120	V
C122 PART	U	F	A	B	D
R2 VALUE	200	470	1K	2.2K	Ω

FOR HEAT SINK RATINGS SEE C122 SPECIFICATION SHEET NUMBER 150.35 AND APPLICATION NOTE NUMBER 200.55

NO HEAT SINK RATINGS AT $T_A \leq 50°$

I CONTACT, MAX.	PULSE WIDTH	DUTY CYCLE
0.67 A	D.C.	100%
4.0 A	160 msec.	12%
8.0 A	160 msec.	3%
12 A	160 msec.	1%
15 A	160 msec.	0.3%

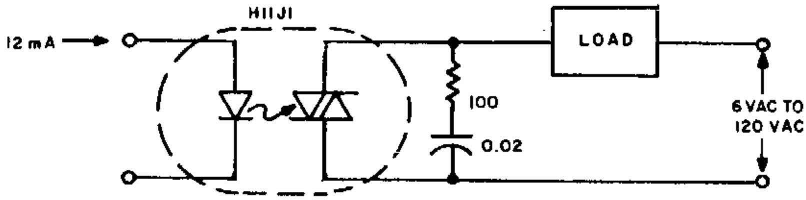
GE

Fig. 67-13

The H11C supplies the dc latching relay function and reverse polarity blocking, for currents up to 300 mA, depending on ambient temperature. For dc use, the gate cathode resistor can be supplemented by a capacitor to minimize transient and *dV/dt* sensitivity. For pulsating dc operation, the capacitor value must be designated to either retrigger the SCR at the application of the next pulse or prevent retriggering at the next power pulse. If not, random or undesired operation might occur. For higher current contacts, the H11C can be used to trigger an SCR capable of handling the current, as illustrated.

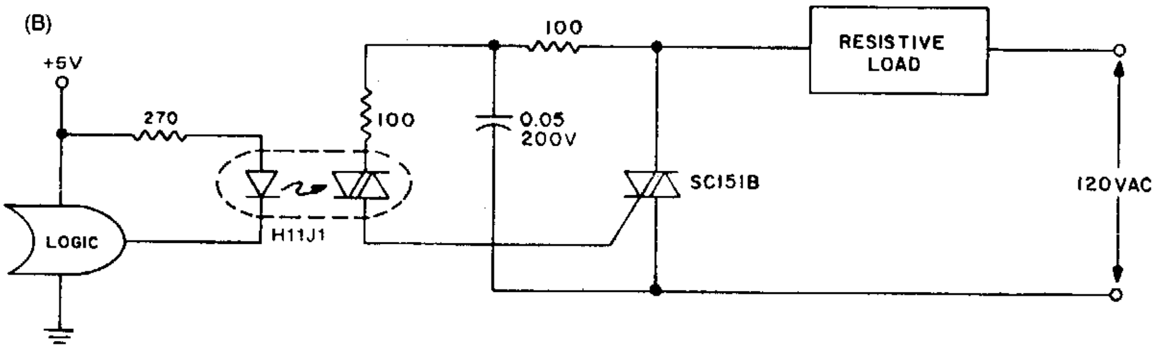
AC RELAY

(A)



SIMPLE "SOLID STATE AC RELAY"

(B)



MINIMUM PARTS COUNT ISOLATED LOGIC TRIGGERED TRIAC

GE

Fig. 67-14

When zero voltage switching is not required, methods of providing this function are illustrated. The lowest parts count version of a solid-state relay is an optoisolator, the triac driver H11J. Unfortunately, the ability of the H11J to drive a load on a 60-Hz line is severely limited by its power dissipation and the dynamic characteristics of the detector. These factors limit applications to 30 – 50 mA resistive loads on 120 Vac, and slightly higher values at lower voltages. These values are compatible with neon lamp drive, pilot, and indicator incandescent bulbs; low voltage control circuits, such as furnace and bell circuits, if *dV/dt* are sufficient; but less than benign loads require a discrete triac.

The H11J1 triac trigger optocoupler potentially allows a simple power switching circuit utilizing only the triac, a resistor, and the optocoupler. This configuration will be sensitive to high values of *dV/dt* and noise on normal power-line voltages, leading to the need for the configuration shown in Fig. 67-14B, where the triac snubber acts as a filter for line voltage to the optocoupler.

Since the snubber is not usually used for resistive loads, the cost effectiveness of the circuit is compromised somewhat. Even with this disadvantage, the labor, board space, and inventory of parts savings of this circuit prove it cost-optimized for isolated logic control of power-line switching. In applications where transient voltages on the power line are prevalent, provisions should be made to protect the H11J1 from breakover triggering.

PHOTODIODE SOURCE FOLLOWER

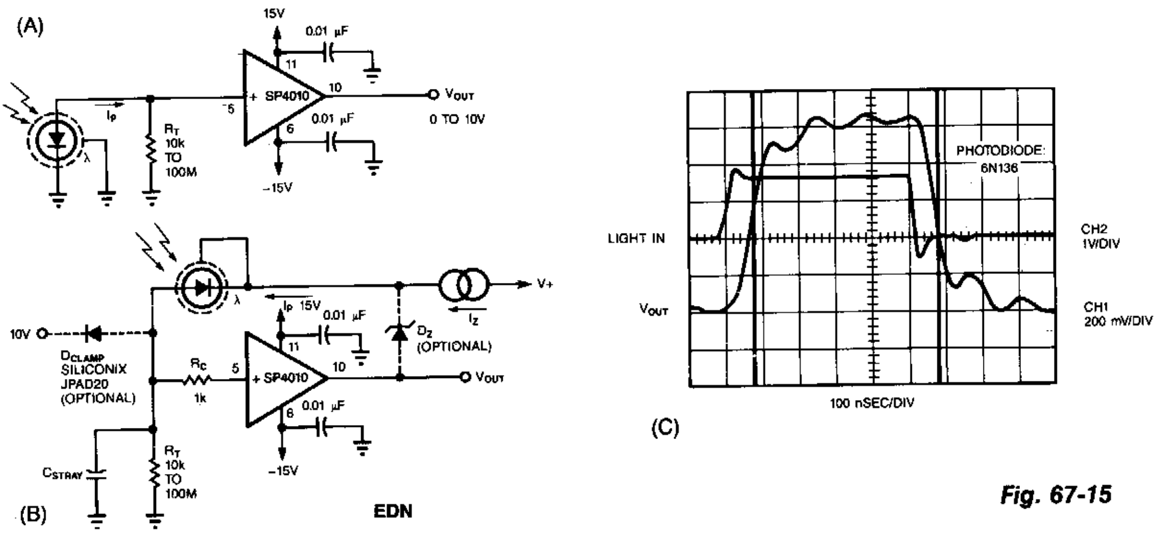


Fig. 67-15

A common method of transforming the output current of a photodiode into a voltage signal, paralleling the photodiode with a high-value load resistor, produces a nonlinear response. Also the combination of the load's transresistance, R_T, and the photodiode's junction capacitance, C_J, slows the circuit's response time. Figure 67-11B shows virtually the same components as Fig. 67-11A rearranged to maximize the inherent speed and linearity of the photodiode. The SP4010 (available from Hybrid Systems, Billerica, MA) is a unity voltage-gain buffer with a JFET input, 60-MHz 3-dB bandwidth, and 18-bit, 0.0004%, linearity over a ± 10 V input range.

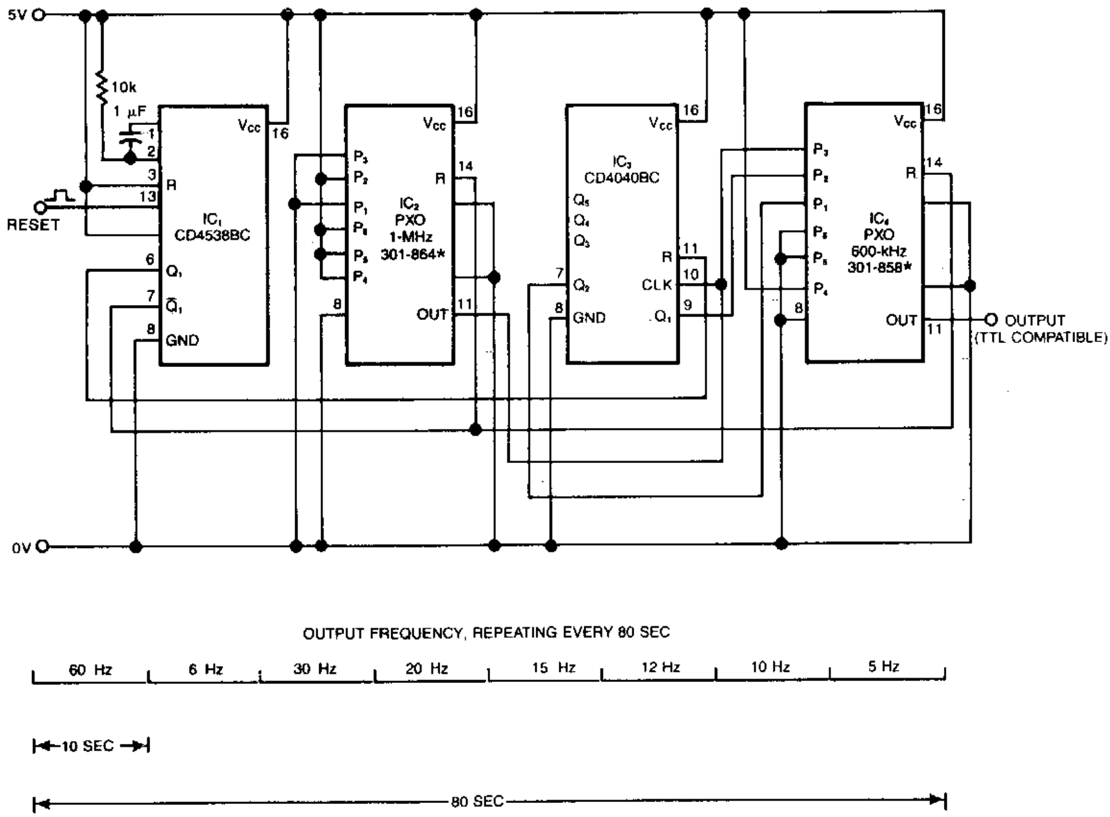
In the circuit of Fig. 67-11B, the photodiode sees a constant voltage across its terminals, which is essential for linear photodiode outputs. The optional zener diode, D_Z, sets a reverse bias at the photodiode for lower junction capacitance and higher speed. If you don't use D_Z, be sure to connect the feedback loop. An optional diode, D_{CLAMP}, limits the output in case of unexpected light bursts, but results in increased dark-current leakage and lower speed. The buffered output of the circuit equals the photodiode current times the transresistance, R_T. Figure 67-11C shows the circuit's response to a fast light pulse.

68

Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DISCRETE SEQUENCE OSCILLATOR



*INTERFACE QUARTZ DEVICES LTD, CREWKERNE, SOMERSET, UK

EDN

Fig. 68-1

The swept-frequency oscillator offers an inexpensive source of discrete frequencies for use in testing digital circuits. In this configuration, the circuit generates an 80-second sequence of eight frequencies, dwelling for 10 seconds on each frequency. You can change the dwell time or the number of frequencies. Frequencies can range from 0.005 Hz to 1 MHz.

The programmable crystal oscillators, PXOs, IC2 and IC4 can each generate 57 frequencies in response to an 8-bit external code. IC2 contains a 1-MHz crystal and produces a 0.05-Hz output. IC4 contains a 600-kHz crystal; its output changes in response to the combined outputs of the 12-stage binary counter IC3 (Q1 and Q2) and the PXO IC2.

To generate more frequencies, you can use one or more of IC3's outputs, (Q3, Q4, Q5) to drive one or more of IC4's inputs (P4, P5, P6). Similarly, you can rewire IC2 or drive it with other logic to control the duration of each frequency. IC1, a monostable multivibrator, provides a system reset. It initiates the sequence shown, beginning at 60 Hz, in response to a positive pulse.

FIXED-FREQUENCY VARIABLE DUTY-CYCLE OSCILLATOR

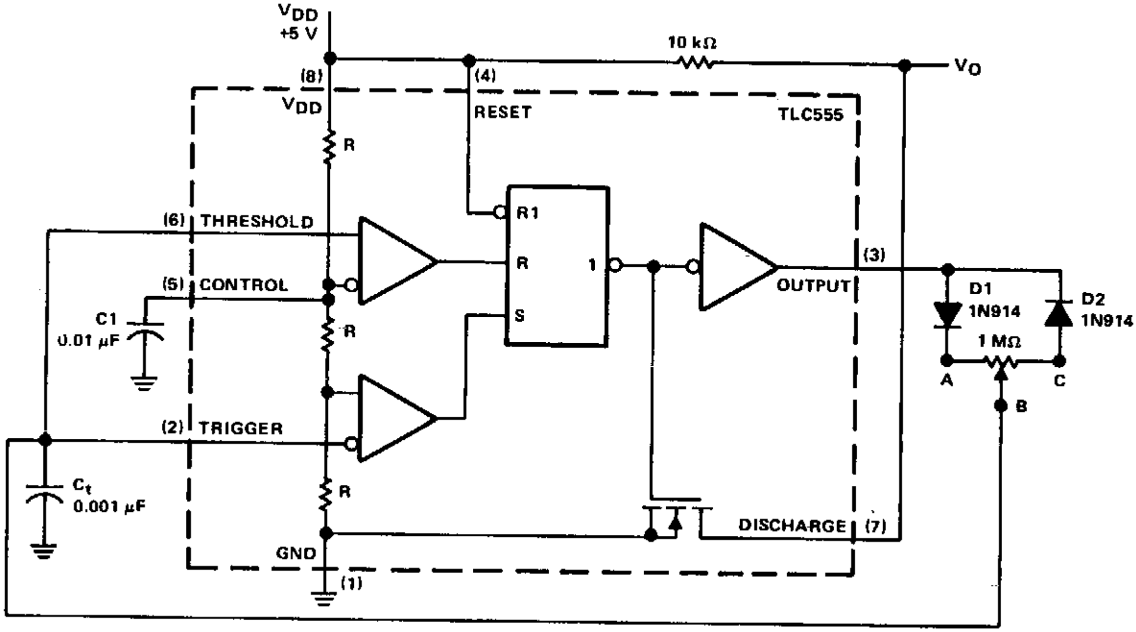
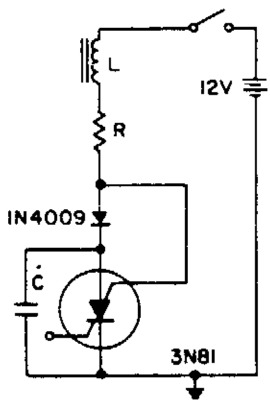


Fig. 68-2

In a basic astable timer, configuration timing periods t_1 and t_2 are not controlled independently. The lack of control makes it difficult to maintain a constant period, T, if either t_1 or t_2 is varied. In this circuit, charge R_{AB} and discharge R_{BC} resistances are determined by the position of common wiper arm B of the potentiometer. So, it is possible to adjust the duty-cycle by adjusting t_1 and t_2 proportionately, without changing period T.

At start-up, the voltage across C_t is less than the trigger level voltage ($1/2\ V_{DD}$), causing the timer to be triggered via pin 2. The output of the timer at pin 3 increases, turning off the discharge transistor at pin 7 and allowing C_t to charge through diode D1 and resistance R_{AB}. When capacitor C_t charges to upper threshold voltage $2/3\ V_{DD}$, the flip-flop is reset and the output at pin 3 decreases. Capacitor C_t then discharges through diode D2 and resistor R_{BC}. When the voltage at pin 2 reaches $1/3\ V_{DD}$, the lower threshold or trigger level, the timer triggers again and the cycle is repeated. In this circuit, the oscillator frequency remains fixed and the duty cycle is adjustable from less than 0.5% to greater than 99.5%.

RLC OSCILLATOR

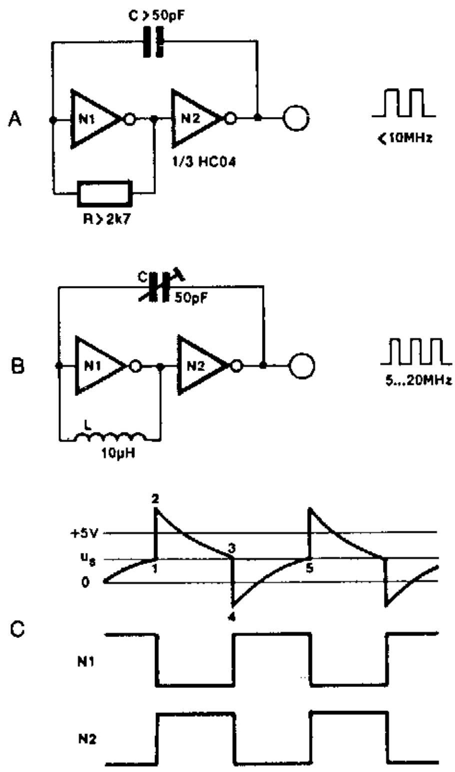


GE

Fig. 68-3

A positive transient, such as the power switch closing, charges C through L to a voltage above the supply voltage, if Q is sufficient. When the current reverses, the diode blocks and triggers the SCS. As the capacitor discharges, the anode gate approaches ground potential, depriving the anode of holding current. This turns off the SCS, and C charges to repeat the cycle.

HC-BASED OSCILLATORS



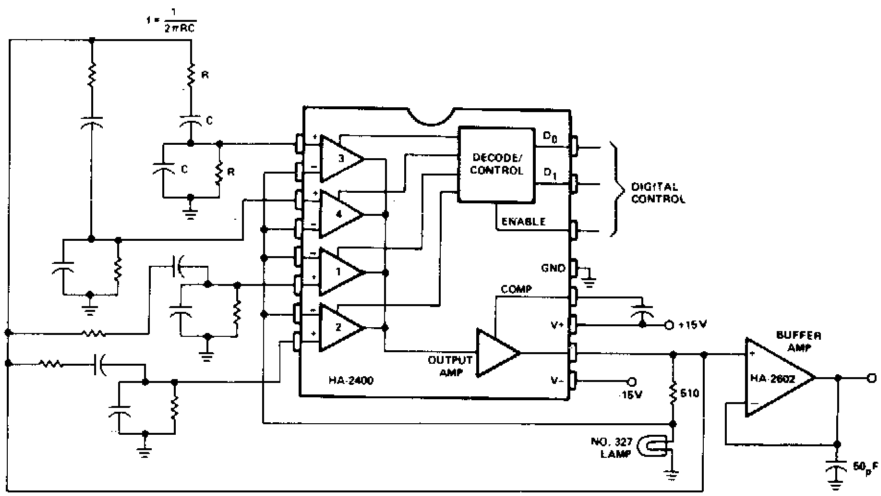
ELEKTOR ELECTRONICS

Fig. 68-4

Two inverters, one resistor, and one capacitor are all that is required to make a HC(T)-based oscillator that gives reliable operation up to about 10 MHz. The use of two HC inverters produces a fairly symmetrical rectangular output signal. In the same circuit, HCT inverters give a duty factor of about 25%, rather than about 50%, since the toggle point of an HC and an HCT inverter is $1/2\ V_{CC}$, and slightly less than 2 V, respectively. If the oscillator is to operate above 10 MHz, the resistor is replaced with a small inductor, as shown in Fig. 68-4B.

The output frequency of the circuit in Fig. 68-4A is given as about 1/1.8rc, and can be made variable by connecting a 100-KΩ potentiometer in series with R. The solution adopted for the oscillator in Fig. 68-4B is even simpler: C is a 50-pF trimmer capacitor.

PROGRAMMABLE-FREQUENCY SINE-WAVE OSCILLATOR

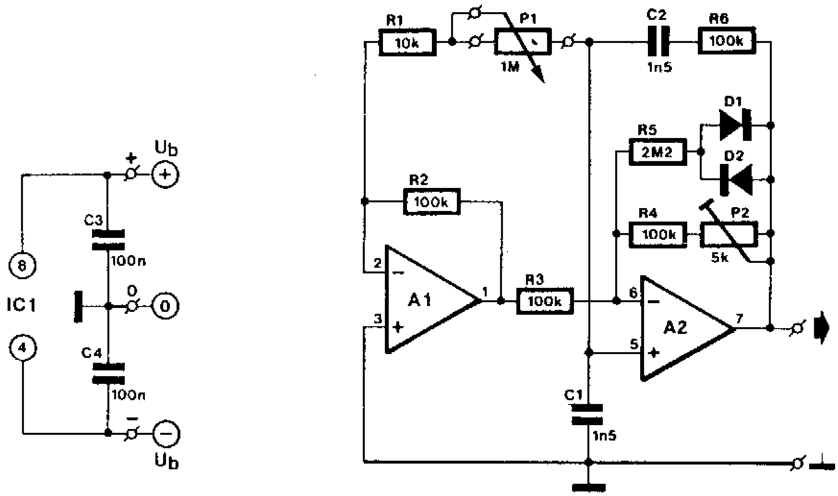


HARRIS

Fig. 68-5

This Wien-bridge oscillator is very popular for signal generators, since it is easily turned over a wide frequency range, and has a very low distortion sine-wave output. The frequency determining networks can be designed from about 10 Hz to greater than 1 MHz; the output level is about 6.0-V rms. By substituting a programmable attenuator for the buffer amplifier, a very versatile sine-wave source for automatic testing, etc. can be constructed.

VARIABLE WIEN-BRIDGE OSCILLATOR



D1, D2 = 1N4148

A1, A2 = IC1 = TLC272, TL072, OP-221 87441

ELEKTOR ELECTRONICS

Fig. 68-6

A Wien-bridge oscillator can be made variable by using two frequency-determining parts that are varied simultaneously at high tracking accuracy. High-quality tracking potentiometers or variable capacitors are, however, expensive and difficult to obtain. To avoid having to use such a component, this oscillator was designed to operate with a single potentiometer. The output frequency, f_o, is calculated from:

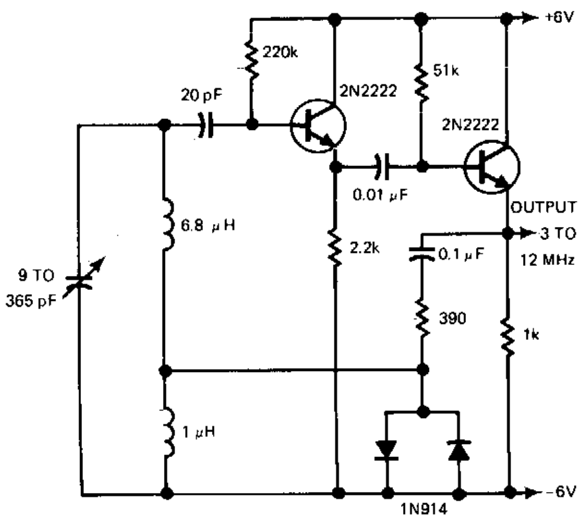
$$f_o = 1/(2nRC\ v\ a)$$

where:

$$R = R_2 = R_3 = R_4 = R_6,\ C = C_1 = C_2,\ \text{and}\ a = (P_1 + R_1)R$$

With preset P2 you can adjust the overall amplification so that the output signal has a reasonably stable amplitude, 3.5 V_{PP} max., over the entire frequency range. The stated components allow the frequency to be adjusted between 350 Hz and 3.5 kHz.

WIDE-RANGE OSCILLATOR



EDN

Fig. 68-7

The gain control allows the oscillator to maintain essentially constant output over its range. The circuit functions over 160 kHz to 12 MHz with essentially constant amplitude.

HCU/HCT-BASED OSCILLATOR

ELEKTOR ELECTRONICS

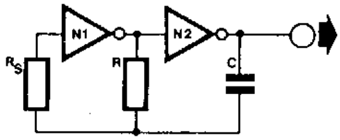


Fig. 68-8

N1, N2 = 1/3 IC1 = 74HCT04, 74HCU04

When frequency stability is not of prime importance, a simple, yet reliable, digital clock oscillator can be made with the aid of relatively few components. High-speed CMOS (HCU/HCT) inverters or gates with an inverter function are eminently suitable to make such oscillators, thanks to their low power consumption, good output signal definition, and extensive frequency range.

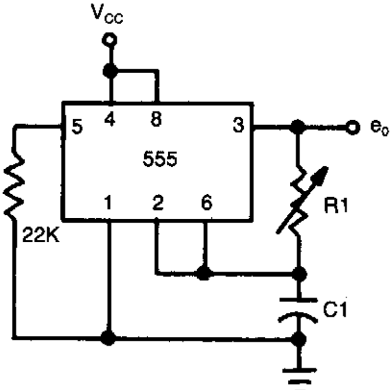
The circuit as shown uses two inverters in a 74HCT04 or 74HCU04. The basic design equations are:

For HCU: $f=1/T$; $T=2.2\,RC$; V_{CC} 6 V; $I_c=13$ mA

For HCT: $f=1/T$; $T=2.4\,RC$; V_{CC} 5.5 V; $I_c=2.2$ mA

With R_s and R calculated for a given frequency and value of C, both resistors can be realized as presets to enable precise setting of the output frequency and the duty factor. Do not forget, however, to fit small series resistors in series with the presets, in observance of the minimum values for R and R_s as given in the design equations. The values quoted for I_c are only valid if the inputs of the remaining gates are grounded.

50% DUTY-CYCLE OSCILLATOR

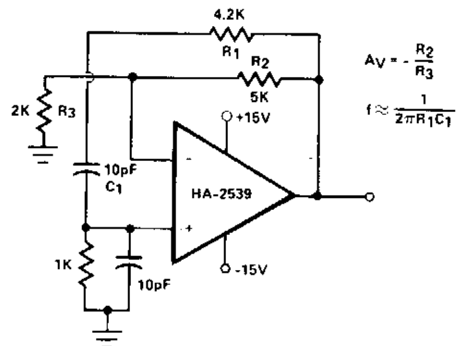


EDN

Fig. 68-9

Frequency of oscillation depends on the R1/C1 time constant and allows frequency adjustment by varying R1.

HIGH-FREQUENCY OSCILLATOR

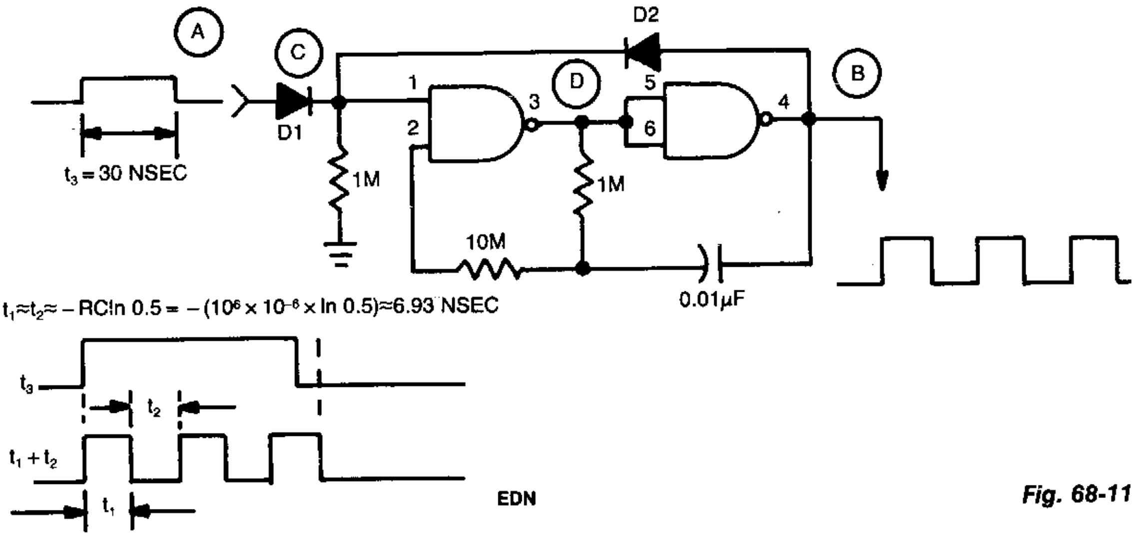


HARRIS

Fig. 68-10

Intended primarily as a building block for a QRP transmitter, this 20-MHz oscillator delivered a clean 6-V, pk-pk signal into a 100-Ω load.

LAST-CYCLE COMPLETING GATED OSCILLATOR

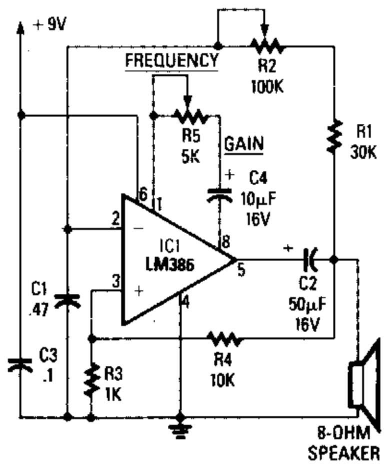


EDN

Fig. 68-11

Regenerative feedback at C enables the oscillator to complete its timing cycle, rather than immediately shutting it off. The IC used was a CD4011AE, although an equivalent will work.

AUDIO OSCILLATOR



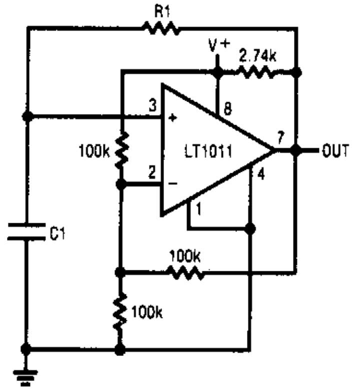
Reprinted with permission of Radio-Electronics Magazine, August 1986.

Fig. 68-12

The circuit's frequency of oscillation is $f = 2.8/[C_1 \times (R_1 + R_2)]$. Using the values shown, the output frequency can be varied from 60 Hz to 20 kHz by rotating potentiometer R2.

A portion of IC1's output voltage is fed to its noninverting input at pin 3. The voltage serves as a reference for capacitor C1, which is connected to the noninverting input at pin 2 of the IC. That capacitor continually charges and discharges around the reference voltage, and the result is a square-wave output. Capacitor C2 decouples the output.

LOW-FREQUENCY OSCILLATOR



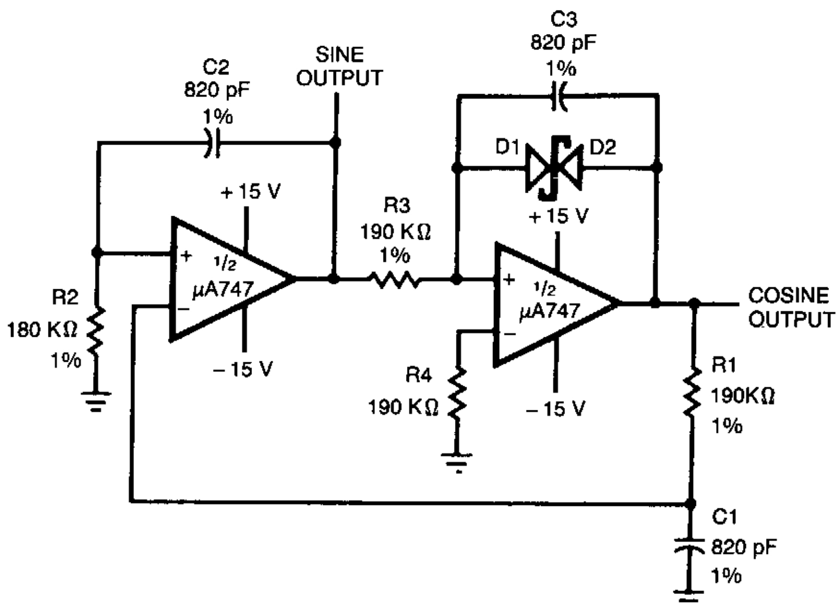
LINEAR TECHNOLOGY CORP.

Fig. 68-13

This simple rc oscillator uses a medium-speed comparator with hysteresis and feedback through R1 and C1 as timing elements. The frequency of oscillation is, at least theoretically, independent from the power supply voltage. If the comparator swings to the supply rails, if the pull-up resistor is much smaller than the resistor R_h, and if the propagation delay is negligible compared to the rc time constant, the oscillation frequency is:

$$f_{OSC} = \frac{0.72}{R_1 C_1}$$

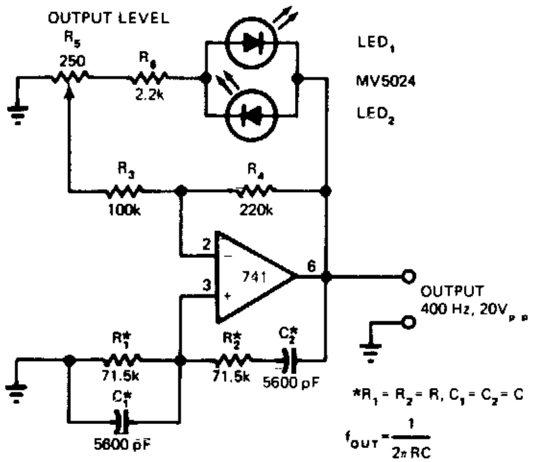
QUADRATURE OSCILLATOR



FAIRCHILD CAMERA AND INSTRUMENT

Fig. 68-14

WIEN-BRIDGE OSCILLATOR

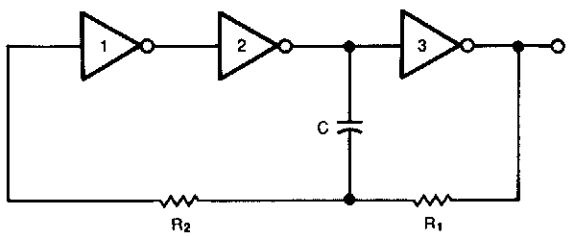


EDN

Fig. 68-15

LEDs function as both pilot lamps and as an AGC (automatic gain control) in this unconventional amplitude-stabilized oscillator.

CMOS OSCILLATOR

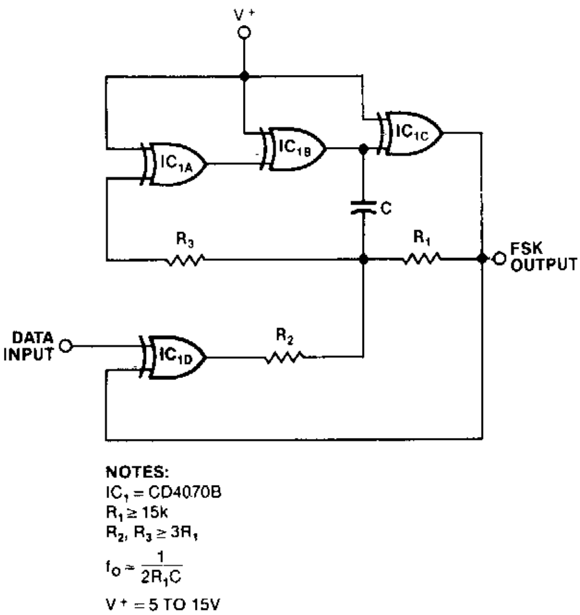


EDN

Fig. 68-16

This circuit is guaranteed to oscillate at a frequency of about $2.2/(R_1 \times C)$ if R_2 is greater than R_1. You can reduce the number of gates further if you replace gates 1 and 2 with a noninverting gate.

XOR-GATE OSCILLATOR

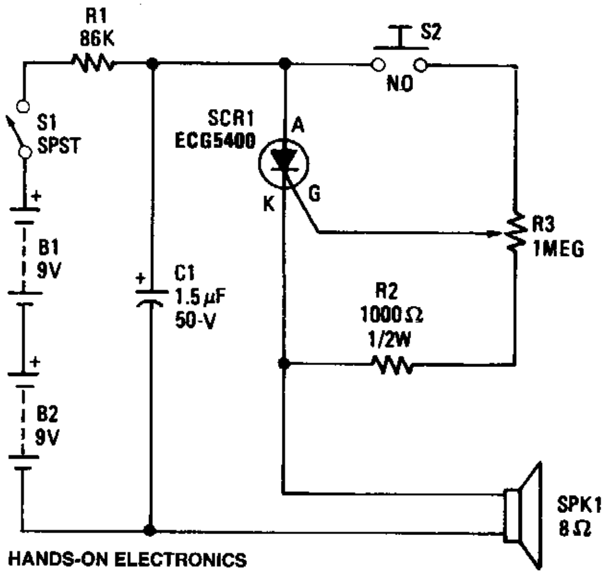


EDN

Fig. 68-17

An exclusive-OR gate, IC1D, turns a simple CMOS oscillator into an FSK generator. When the data input increases, IC1D inverts, and negative feedback through R2 lowers the circuit's output frequency. A low input results in positive feedback and a higher output frequency. R1 and C set the oscillator's frequency range, and R2 determines the circuit's frequency shift. To ensure frequency stability, make R3 much greater than R1 and use a high-quality feedback capacitor. The three gates constituting the oscillator itself need not be exclusive-OR types; use any CMOS inverter.

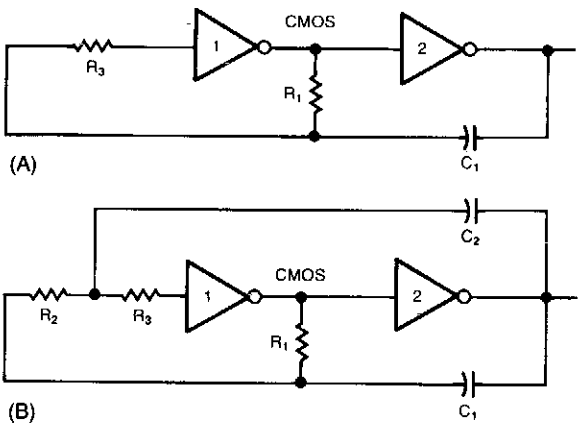
SCR RELAXATION OSCILLATOR



HANDS-ON ELECTRONICS

Fig. 68-18

CMOS OSCILLATOR



EDN

Fig. 68-19

The common clock oscillator in Fig. 68-19A has two small problems: It might not, in fact, oscillate if the transition regions of its two gates differ. If it does oscillate, it might sometimes oscillate at a slightly lower frequency than its equation predicts because of the finite gain of the first gate. If the circuit does work, oscillation occurs usually because both gates are in the package and, therefore, have logic thresholds only a few millivolts apart.

The circuit in Fig. 68-19B resolves both problems by adding a resistor and a capacitor. The R2/C2 network provides hysteresis, thus delaying the onset of gate 1's transition until C1 has enough voltage to move gate 1 securely through its transition region. When gate 1 is finally in its transition region, C2 provides positive feedback, thus rapidly moving gate 1 out of its transition region.

The equations for the oscillator in Fig. 68-19B are:

$$R_2 = 10R_1$$
$$R_3 = 10R_2$$
$$C_1 = 100C_2$$

$$f = \cong \frac{1}{1.2R_1C_1}$$

CODE-PRACTICE OSCILLATOR

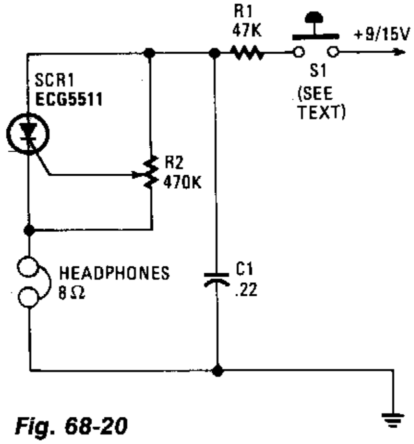
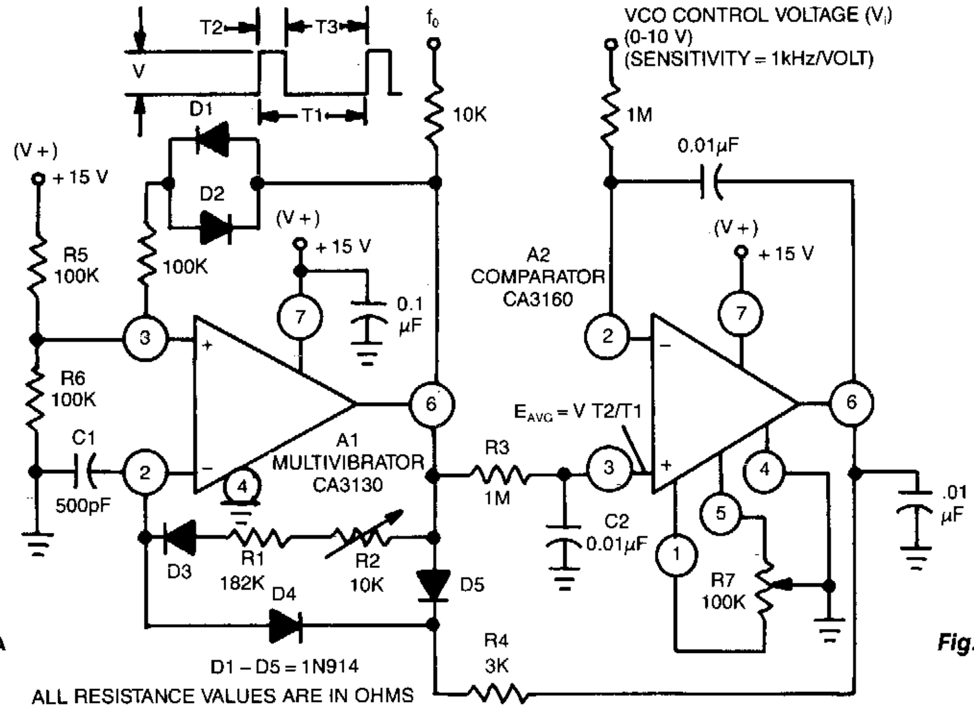


Fig. 68-20

Capacitor C1 charges through resistor R1, and when the gate level established by potentiometer R2 is high enough, the SCR is triggered. Current flows through the SCR and earphones, discharging C1. The anode voltage and current drop to a low level, so the SCR stops conducting and the cycle is repeated. Resistor R2 lets the gate potential across C1 be adjusted, which charges the frequency or tone. Use a pair of 8-Ω headphones. The telegraph key goes right into the B+ line, 9-V battery.

**HANDS-ON ELECTRONICS/
POPULAR ELECTRONICS**

PRECISION VOLTAGE-CONTROLLED OSCILLATOR

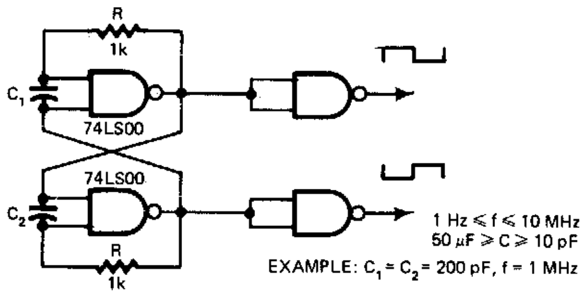


GE/RCA

Fig. 68-21

This circuit uses a CA3130 BiMOS op amp as a multivibrator and CA3160 BiMOS op amp as a comparator. The oscillator has a sensitivity of 1 kHz/V, with a tracking error in the order of 0.02%, and a temperature coefficient of 0.01%/°C.

5-V OSCILLATOR

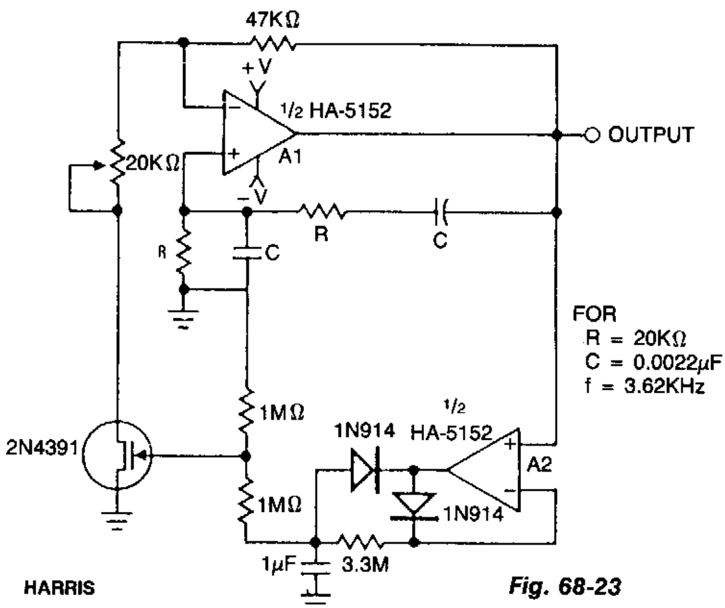


EDN

Fig. 68-22

Consistently self-starting and yet capable of operating from over 1 Hz to 10 MHz, this low-cost oscillator requires only five components. Calculate the period of oscillation by using this relationship: $P = 5 \times 10^3\, C$ sec when $C = C_1 = C_2$. By changing the ratio of C1 to C2, the duty cycle can be as low as 20%.

LOW-VOLTAGE WIEN-BRIDGE OSCILLATOR



HARRIS

Fig. 68-23

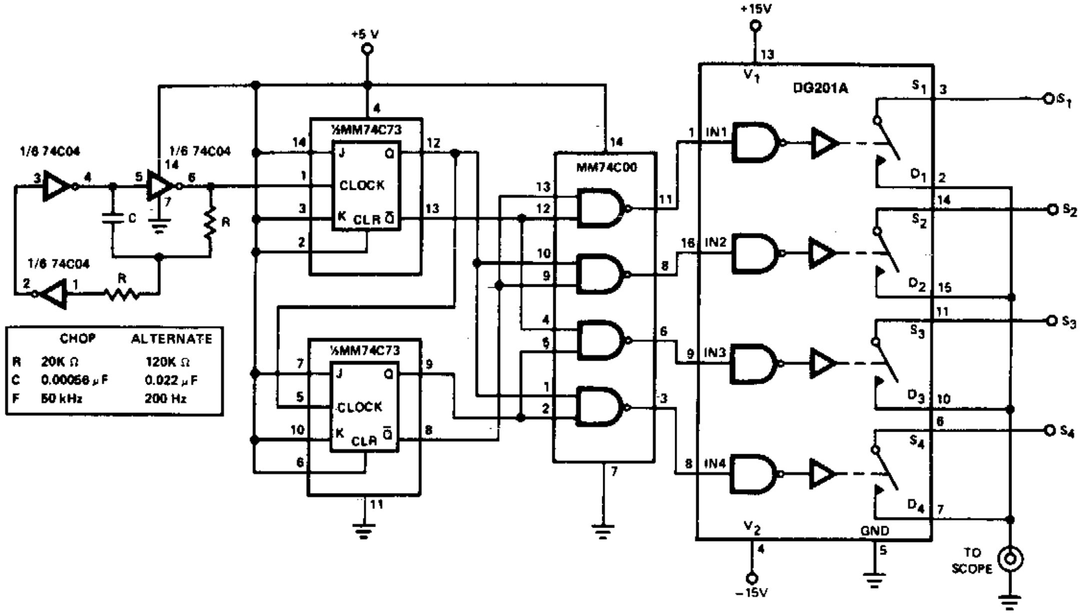
This circuit utilizes an HA-5152 dual op amp and FET to produce a low-voltage, low-power, Wein-bridge sine-wave oscillator. Resistors *R* and capacitors *C* control the frequency of oscillation; the FET, used as a voltage-controlled resistor, maintains the gain of A1 exactly 3 dB to sustain oscillation. The 20-KΩ pot can be used to vary the signal amplitude. The HA-5152 has the capability to operate from ±1.5-V supplies. This circuit will produce a low-distortion sine-wave output while drawing only 400 uA of supply current.

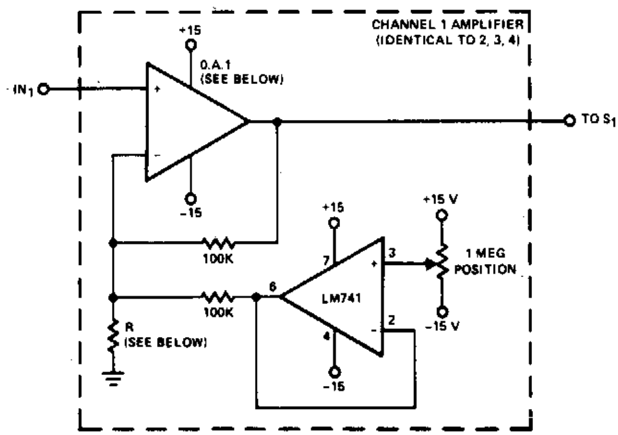
69

Oscilloscope Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SCOPE EXTENDER





O.A.1 IS OP AMP WITH SUITABLE BANDWIDTH, SLEW RATE, ETC., FOR DESIRED SIGNALS

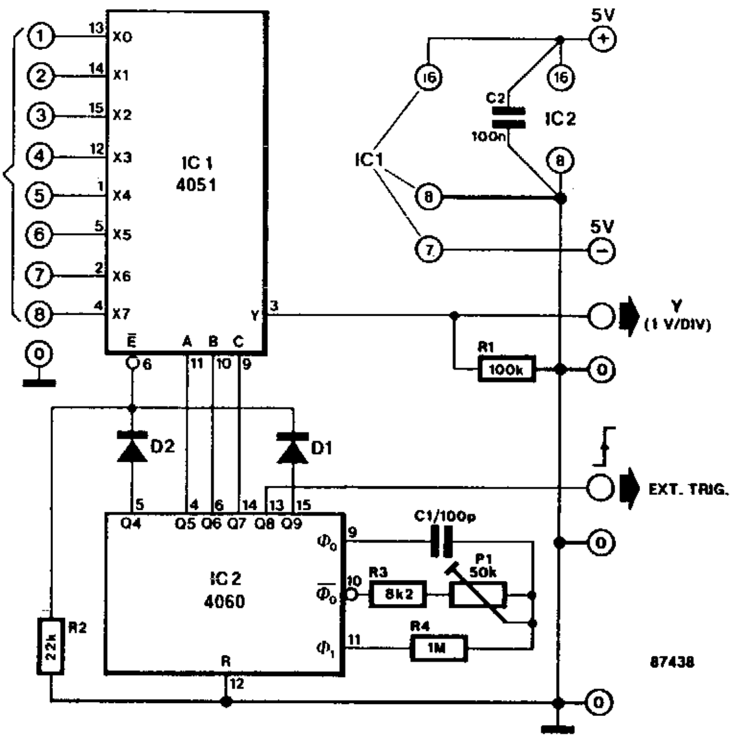
R IS ADDED FOR EXTRA GAIN ACCORDING TO FORMULA VOLTAGE GAIN = $2 + \frac{100K}{R}$

SILICONIX

Fig. 69-1

The adapter allows four inputs to be displayed simultaneously on a single trace scope. For low-frequency signals, less than 500 Hz, the adapter is used in the *chop* mode at a frequency of 50 kHz. The clock can be run faster, but switching glitches and the actual switching time of the DG201A limit the maximum frequency to 200 kHz. High frequencies are best viewed in the alternate mode, with a clock frequency of 200 Hz. When the clock is below 100 Hz, trace flicker becomes objectionable. One of the four inputs is used to trigger the horizontal trace of the scope.

EIGHT-CHANNEL VOLTAGE DISPLAY



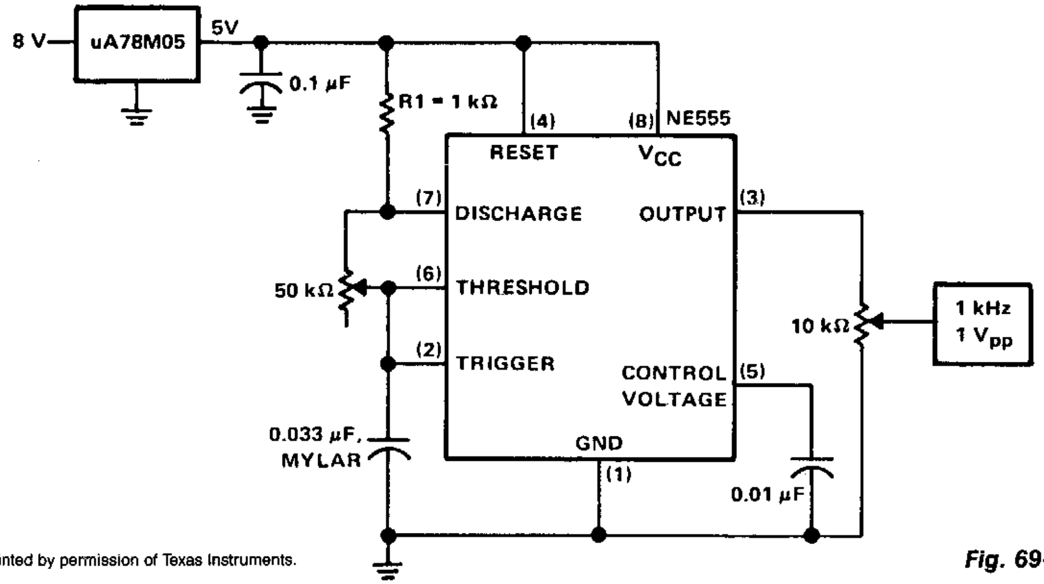
ELEKTOR ELECTRONICS *Fig. 69-2*

This circuit turns a common oscilloscope into a versatile eight-channel display for direct voltages. The trend of each of the eight input levels is readily observed, albeit that the attainable resolution is not very high.

The circuit diagram shows the use of an eight-channel analog multiplexer IC1, which is the electronic version of an eight-way rotary switch with contacts X0 through X7 and pole Y. The relevant channel is selected by applying a binary code to the A-B-C inputs. For example, binary code 011 (A-B-C) enables channel 7 (X6 Y). The A-B-C inputs of IC1 are driven from three successive outputs of binary counter IC2, which is set to oscillate at about 50 kHz with the aid of P1. Since the counter is not reset, the binary state of outputs Q5, Q6, and Q7 steps from 0 to 7 in a cyclic manner. Each of the direct voltages at input terminals 1 to 8 is therefore briefly connected to the Y input of the oscilloscope. All eight input levels can be seen simultaneously by setting the timebase of the scope, in accordance with the time it takes the counter to output states 0 through 7, on outputs Q5, Q6, and Q7.

The timebase on the scope should be set to 0.5 ms/div, and triggering should occur on the positive edge of the external signal. Set the vertical sensitivity to 1 V/div. The input range of this circuit is from −4 V to +4 V; connected channels are terminated in about 100 KΩ.

OSCILLOSCOPE CALIBRATOR

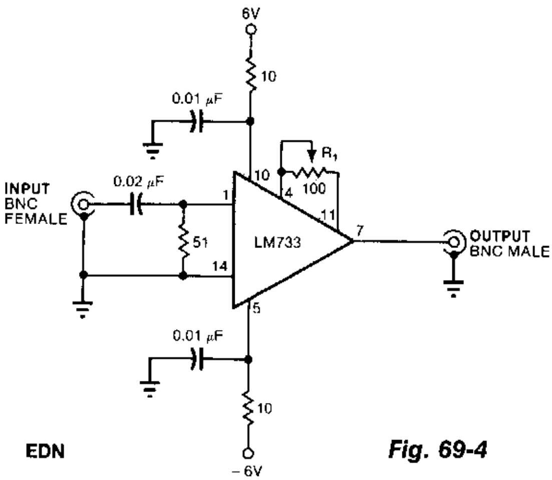


Reprinted by permission of Texas Instruments.

Fig. 69-3

The calibrator can be used to check the accuracy of a time-base generator, as well as to calibrate the input level of amplifiers. The calibrator consists of an NE555 connected in the astable mode. The oscillator is set to exactly 1 kHz by adjusting potentiometer P1 while the output at pin 3 is being monitored against a known frequency standard or frequency counter. The output level, likewise, is monitored from potentiometer P2's center arm to ground with a standard instrument. P2 is adjusted for 1 V pk-pk at the calibrator output terminal. During operation, the calibrator output terminal will produce a 1-kHz, square-wave signal at 1 V pk-pk with about 50% duty cycle. For long-term oscillator frequency stability, C1 should be a low-leakage mylar capacitor.

SCOPE SENSITIVITY AMPLIFIER



EDN

Fig. 69-4

This circuit provides 20 ±0.1 dB voltage gain from 0.5 to 25 MHz and ±3 dB from 70 kHz to 55 MHz. An LM733 video amplifier furnishes a low input-noise spec, 10-μV typical, measured over a 15.7-MHz bandwidth. The scale factor of the instrument can be preserved by using a trimmer R1 or a selected precision resistor, to set the circuit's voltage gain to exactly 100.

ADD-ON SCOPE MULTIPLEXER

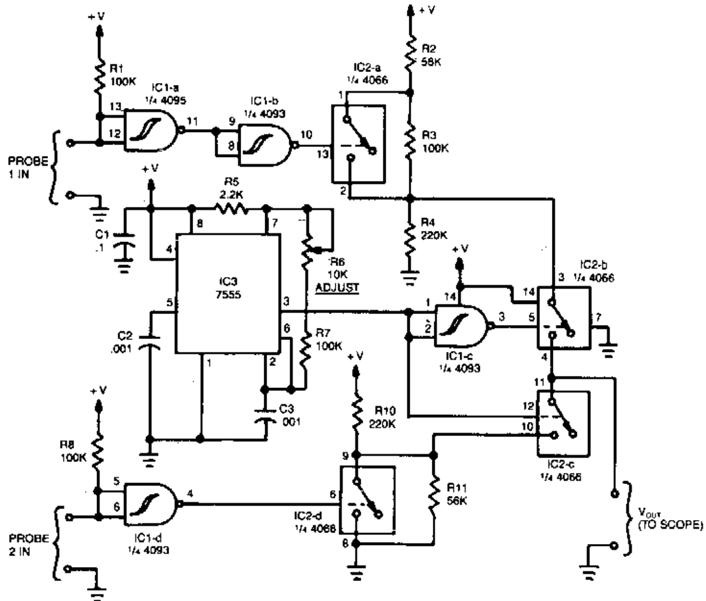
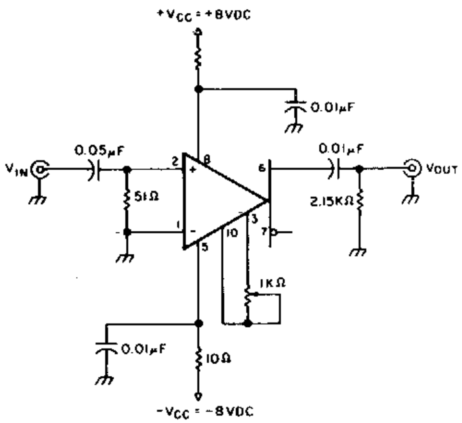


Fig. 69-5

The operation of the unit revolves around three ICs: a 4093 quad NAND Schmitt trigger, a 4066 quad analog switch, and a 7555 timer. When a high is fed to probe 1 in, it is inverted to IC1a and once again by IC1b, so that the input to IC2a is high. That high causes the *switch contacts* in IC2a to close. With the *contacts* closed, a high-level output is presented to the input of IC2b. The high output is fed to probe 2 in. That signal is then inverted by IC1d and routed to IC2d, causing its *contacts* to open, and the unit to output a logic-level high. The output of IC2d is then fed to IC2c.

OSCILLOSCOPE PREAMPLIFIER



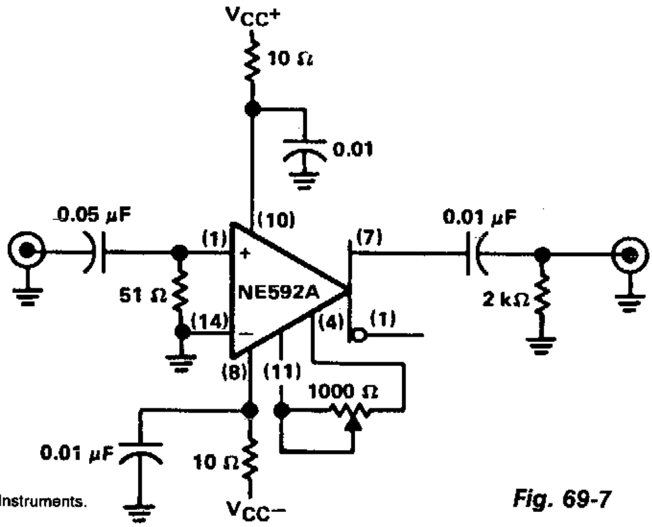
73 AMATEUR RADIO

Fig. 69-6

This circuit provides about 20 dB voltage gain with a frequency range from 0.5 to 50 MHz. You can extend the low-frequency response of this circuit by increasing the value of the 0.05-μF capacitor—or try removing the capacitor. This circuit delivers a particularly small level of input noise, measured at approximately 20 μV over a bandwidth range of 15 MHz.

Calibrate the gain by adjusting the gain potentiometer connected between pins 3 and 10, then adjust the 1-KΩ trimmer potentiometer for an exact voltage gain of 10; this helps preserve the scale factor of the oscilloscope.

OSCILLOSCOPE/COUNTER PREAMPLIFIER

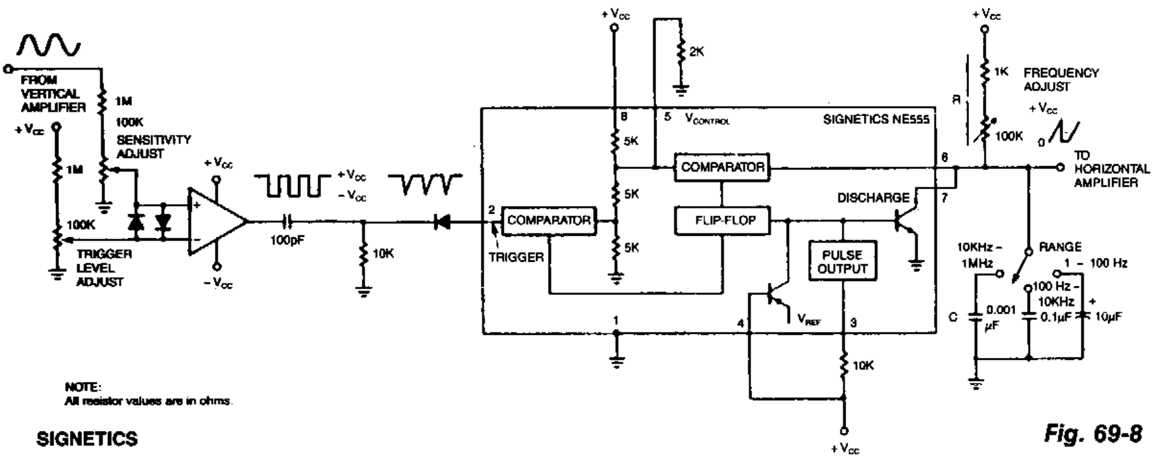


Reprinted by permission of Texas Instruments.

Fig. 69-7

The circuit will provide a 20 ±0.1 dB voltage gain from 500 kHz to 50 MHz. The low-frequency response of the amplifier can be extended by increasing the value of the 0.05-μF capacitor connected in series with the input terminal. This circuit will yield an input-noise level of approximately 10 μV over a 15.7-MHz bandwidth. The gain can be calibrated by adjusting the potentiometer connected between pins 4 and 11. The 1-KΩ potentiometer can be adjusted for an exact voltage gain of 10. This preserves the scale factor of the instrument.

OSCILLOSCOPE-TRIGGERED SWEEP

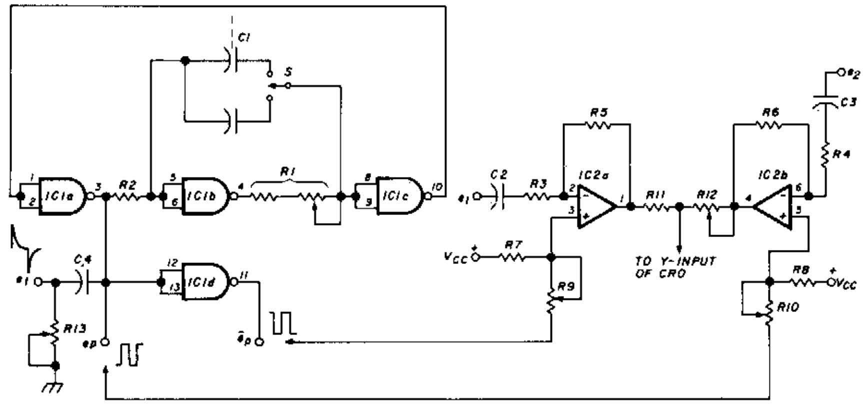


SIGNETICS

Fig. 69-8

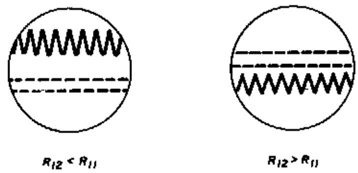
The circuit's input op amp triggers the timer, sets its flip-flop and cuts off its discharge transistor so that capacitor C can charge. When capacitor voltage reaches the timer's control voltage of 0.33 V_{CC}, the flip-flop resets and the transistor conducts, discharging the capacitor. Greater linearity can be achieved by substituting a constant-current source for frequency adjust resistor R.

CRO DOUBLER

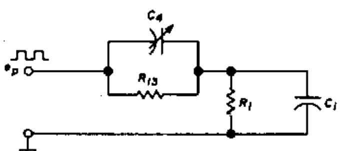


Parts List (fig. 1)

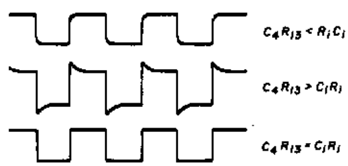
IC1	4011	R3	20k
IC2	LM358	R4	20k
		R5	200k
C1	0.001 μF	R6	200k
	0.01 μF	R7	50k
	0.1 μF	R8	50k
	1.0 μF	R9	50k pot
	10 μF	R10	50k pot
C2	25 μF	R11	100 ohm
C3	25 μF	R12	300 ohm pot
C4	0.001 μF	R13	50k pot
R1	500-ohm + 50-k pot	Vcc	6 volts
R2	500 ohm	S	multipole switch



Oscillograms of the displayed input signals e_1 (sine wave) and e_2 (square wave).



Probe adjusting circuit for CRO.



Displayed square waves on CRO screen.

Fig. 69-9

HAM RADIO

IC1a, IC1b, and IC1c of the quad two-input NAND gate 4011 are connected as an astable multivibrator; IC1d is connected as an inverter. Terminals 3 and 11 of the 4011 produce square waves with opposite phases. The square waves, e_p, at the output of IC1a, passing through differentiator C4R13, then form positive and negative pulses, e_t. The dual op amps of the LM358 are used as two gated amplifiers for singles e_1 and e_2 and fed through terminals 2 and 6, to be displayed simultaneously on the CRO screen.

The two opposite-phase square waves $\bar{e}_p$ and e_p are used to gate IC2a and IC2b at terminals 3 and 5 of the LM358, respectively. Resistances R9 and R10 are preadjusted so that one op amp is driven to saturation while the other works normally as an amplifier. Thus, they will amplify signals e_1 and e_2 alternately, and two separate traces will be displayed on the screen. Resistance R12 can be varied to adjust the vertical separation of the two traces.

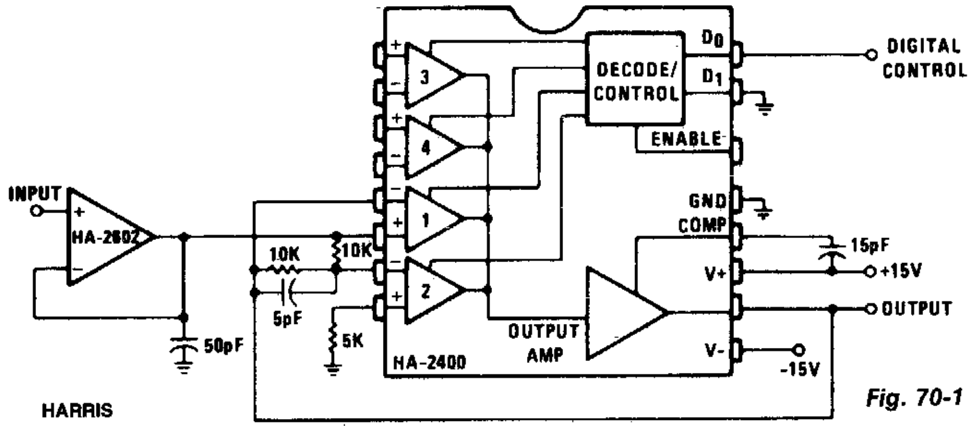
Select a suitable value for C1 with switch S, and adjust the pot of R1. The frequency of square waves can be varied from 1 to 10^6 cps. This process is necessary for stabilizing the waveforms displayed on the screen. A common supply of 6 V is used in the circuit.

70

Phase Detectors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

PHASE SELECTOR/PHASE DETECTOR/ SYNCHRONOUS RECTIFIER/BALANCED MODULATOR

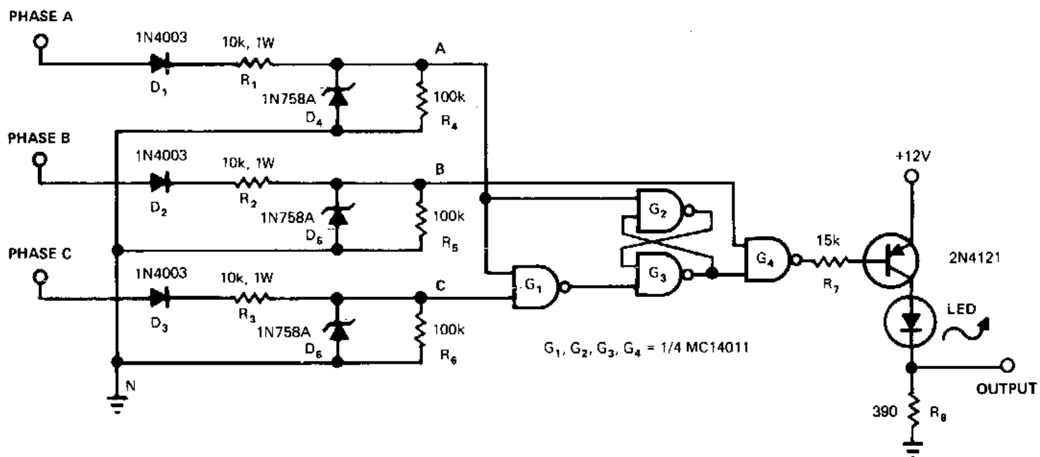


HARRIS

Fig. 70-1

This circuit passes the input signal at unity gain, either unchanged or inverted, depending on the digital control input. A buffered input is shown, since low-source impedance is essential. Gain can be added by modifications to the feedback networks. Signals up to 100 kHz can be handled with 20.0-V pk-pk, output. The circuit becomes a phase detector when driving the digital control input with a reference phase at the same frequency as the input signal; the average dc output is proportional to the phase difference, with 0 V at +90°. By connecting the output to a comparator, which in turn drives the digital control, a synchronous full-wave rectifier is formed. With a low-frequency input signal and a high-frequency digital control signal, a balanced (suppressed carrier) modulator is formed.

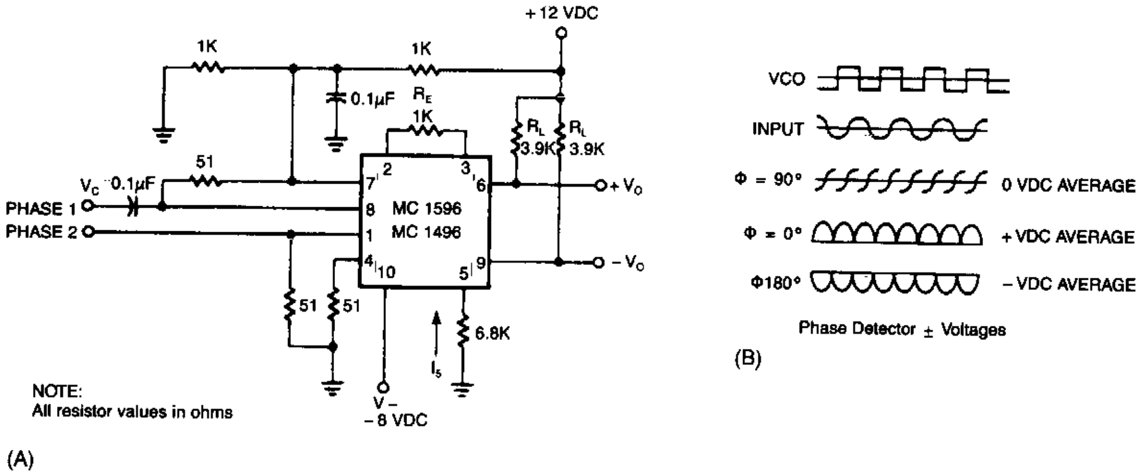
PHASE-SEQUENCE DETECTOR



EDN

Fig. 70-2

PHASE DETECTOR



SIGNETICS

Fig. 70-3

The output of the detector contains a term related to the cosine of the phase angle. Two signals of equal frequency are applied to the inputs. The frequencies are multiplied together, producing the sum and difference frequencies. Equal frequencies cause the difference component to become dc, while the undesired sum component is filtered out. The dc component is related to the phase angle by the graph of Fig. 70-2B. At 90°, the cosine becomes zero, while being at maximum positive or maximum negative at 0° and 180°, respectively. The advantage of using the balanced modulator over other types of phase comparators is the excellent conversion linearity. This configuration also provides a conversion gain, rather than a loss for greater resolution. Used in conjunction with a phase-locked loop, for instance, the balanced modulator provides a very low-distortion FM demodulator.

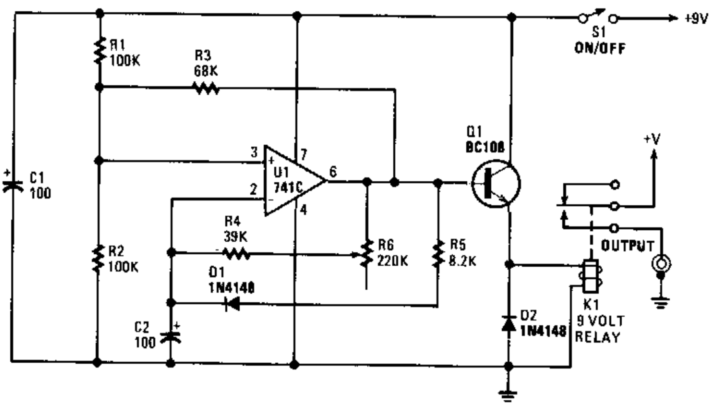
Correct phase sequences (ABC, BCA, or CAB) produce trains of output pulses and illuminate the LED. The output stays low and the LED remains dark for incorrect sequences (BAC, ACB, or CBA) or for phase loss (phase A, B, or C missing).

71

Photography-Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SLIDE-SHOW TIMER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 71-1*

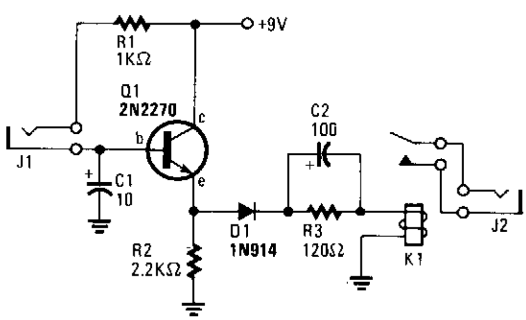
You can set the interval from about 5 – 30 seconds. A relay operates the slide-change mechanism. Op amp U1 forms a sort of Schmitt trigger. Resistors R1 and R2 bias the noninverting input at pin 3 of U1 to half the supply voltage. Feedback resistor R3 increases or reduces the bias to pin 3, depending on whether the output of U1 is high or low.

When power is first applied to the circuit, C2 has a zero charge and the inverting input of the op amp is at a lower voltage than its noninverting input. When the output of U1 is high, C2 begins to charge through R5 and D1. It takes about one second for the charge on C2 to reach the same voltage as that at the noninverting input of U1. At that time, the output of U1 begins a negative swing.

Because of the positive feedback through R3, the voltage at the noninverting input is reduced and the output becomes more negative. The voltage at the noninverting input is about 1/4 of the supply voltage, and C2 begins to discharge through the resistor bank. The timing is controlled by R6.

The resulting pulses are fed to the base of Q1, configured as an emitter-following buffer stage, which is used to activate relay K1. Transistor Q1 is necessary because op amps usually have an output current in the 20-mA range, which is too low to activate the relay.

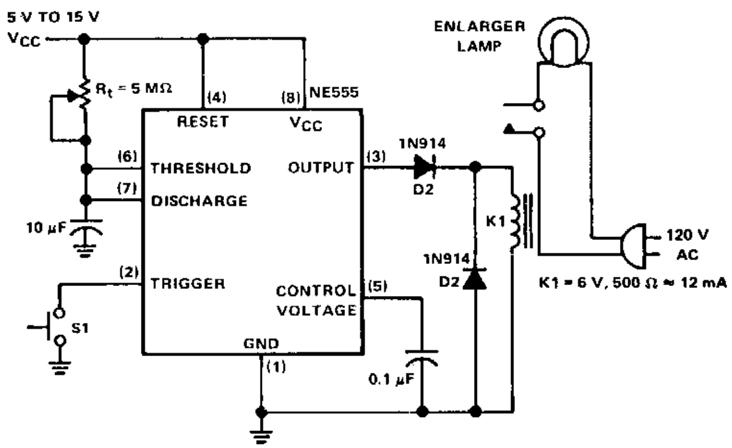
CAMERA ALARM TRIGGER



HANDS-ON ELECTRONICS *Fig. 71-2*

Transistor Q1 remains off until the magnetic switch connected to J1 closes. When that happens, the 9 V is connected through R1 to Q1'S base. Q1 turns on, thereby charging C2 through relay K1, which causes K1's contacts to close. Since the contacts connect via J2 to the remote control jack in the camera, which in turn connects to the camera's shutter release, the closure of K1's contacts will cause the camera's shutter to trigger. After C2 charges, K1 opens because current through its coil ceases; the camera won't take another picture. If Q1 turns off because the magnetic switch on the window or gate opens, C2 discharges and the circuit is ready for another cycle. As long as Q1 remains on, C2 stays charged and prevents K1 from triggering more photos. Capacitor C1 bypasses spurious magnetic switch noises from physical phenomena, such as a rattling window or a gate shaking in the wind, thus reducing the likelihood of an unwanted picture. Resistor R2 biases Q1 and dampens K1/C2 oscillations which might cause contact bounce. Diode D1 prevents C2 from discharging through K1; the relay coil isn't polarity conscious and C2 discharging through it would trigger an unwanted picture.

DARKROOM ENLARGER TIMER



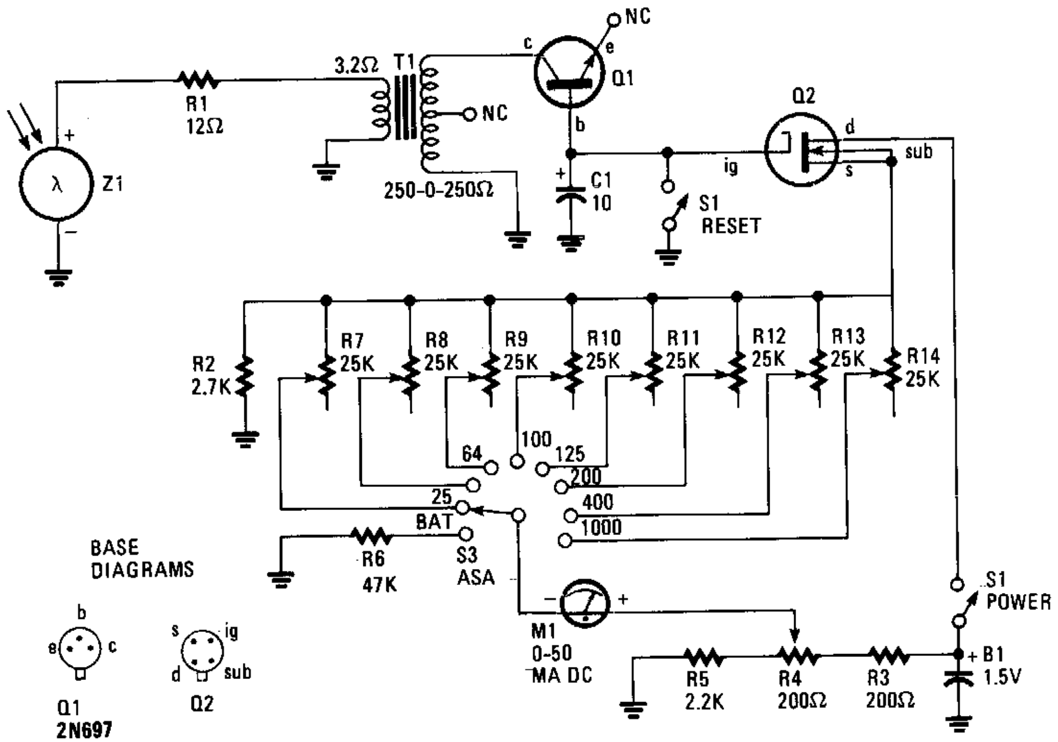
Reprinted by permission of Texas Instruments.

Fig. 71-3

The NE555 circuit is a basic one-shot timer with a relay connected between the output and ground. It is triggered with the normally open momentary contact switch, which when operated, grounds the trigger input at pin 2. This causes a high output to energize K1 which closes the normally open contacts in the lamp circuit. They remain closed during the timing interval, then open at time out. Timing is controlled by a 5-MΩ potentiometer, R_t. All timer-driven relay circuits should use a reverse clamping diode, such as D1, across the coil. The purpose of diode D2 is to prevent a timer output latch-up condition in the presence of reverse spikes across the relay.

With the rc time constant shown, the full-scale time is about 1 minute. A scale for the 5-MΩ potentiometer shaft position can be made and calibrated in seconds. Longer or shorter full-scale times can be achieved by changing the values of the rc timing components.

FLASH METER



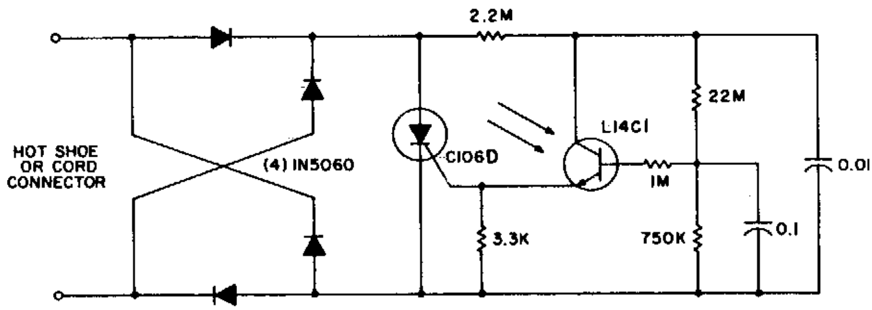
HANDS-ON ELECTRONICS

Fig. 71-4

Insulated-gate, field-effect transistor (IGFET), Q2 and silicon photo cell Z1 form the heart of this circuit. Transformer T1 is an audio-output type, but it's reversed in the circuit. A sudden flash from a photoflash unit detected by Z1 sends a voltage pulse through the low-impedance winding of T1 via R1. That voltage pulse is stepped-up in T1's 500-Ω, primary winding before being rectified by Q1. Transistor Q1 is used as a diode; its emitter lead was snipped off close to the case. Q1 then charges C1 to a value proportional to the amplitude of the electrical pulse generated by the light from a flash unit.

Capacitor C1 controls the current flowing through Q2, which has a very high-input impedance. The current through Q2 is read by meter M1, a 0 – 50 μA dc unit, which has been calibrated in f-stops. The extremely high internal resistances of Q1 and Q2 will allow C1 to retain its charge for several minutes; this is more than enough time for you to take your reading of M1. The charge on C1 is shorted to ground and returned to 0 V by depressing reset button S1. The flashmate is ready to read the next photoflash. Trim potentiometers, R7 through R14, are adjusted to values which will yield correct readings for corresponding film sensitivities, or exposure indexes.

SLAVE PHOTOGRAPHIC XENON FLASH TRIGGER



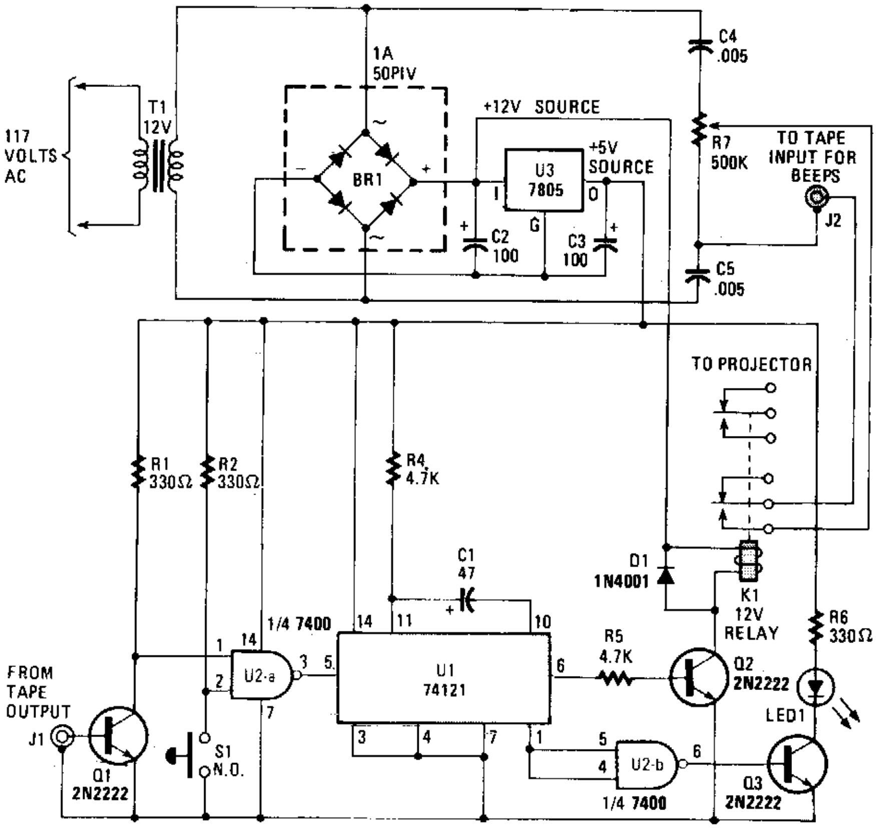
GE

Fig. 71-5

This circuit is used for remote photographic flash units that flash at the same time as the flash attached to the camera. This circuit is designed to the trigger cord or ***hot shoe*** connection of a commercial portable flash unit and triggers the unit from the light produced by the light of the flash unit attached to the camera. This provides remote operation without the need for wires or cables between the various units. The flash trigger unit should be connected to the slave flash before turning the flash on to prevent a dV/dt triggered flash on connection. The L14C1 phototransistor has a wide, almost cosine viewing angle, so alignment is not critical. If a very sensitive, more directional remote trigger unit is desired, the circuit can be modified using an L14G2 lensed phototransistor as the sensor.

The lens on this transistor provides a viewing angle of approximately 10° and gives over a 10 to 1 improvement in light sensitivity (3 to 1 range improvement). Note that the phototransistor is connected in a self-biasing circuit which is relatively insensitive to slow-changing ambient light, and yet discharges the 0.01-μF capacitor into the C106D gate when illuminated by a photo flash. For a physically smaller size, the C106D can be replaced by a C205D, if the duty cycle is reduced appropriately.

SLIDE TIMER



POPULAR ELECTRONICS

Fig. 71-6

This circuit will record commentary and/or music on one track of a tape and put the beeps that change the slides on another track. Gate U2a is used to trigger U1, which is configured as a timer when a pulse is received from either the tape input or via pushbutton switch S1. Timer U2 outputs one pulse for every input pulse received, no matter how long S1 is depressed.

The *Q* output of U2 at pin 1 is fed to U2b, which is set up as an inverter. When pin 1 of U1 becomes low, Q3 is activated, lighting LED1. The *Q* output of U1 at pin 6 is tied to the base of Q2, through R5, so that when pin 6 becomes high, Q2 is turned on. When Q2 is turned on, relay K1 is energized, and a signal is fed to the tape input through J2. The second set of contacts of K1 are used to trigger the projector.

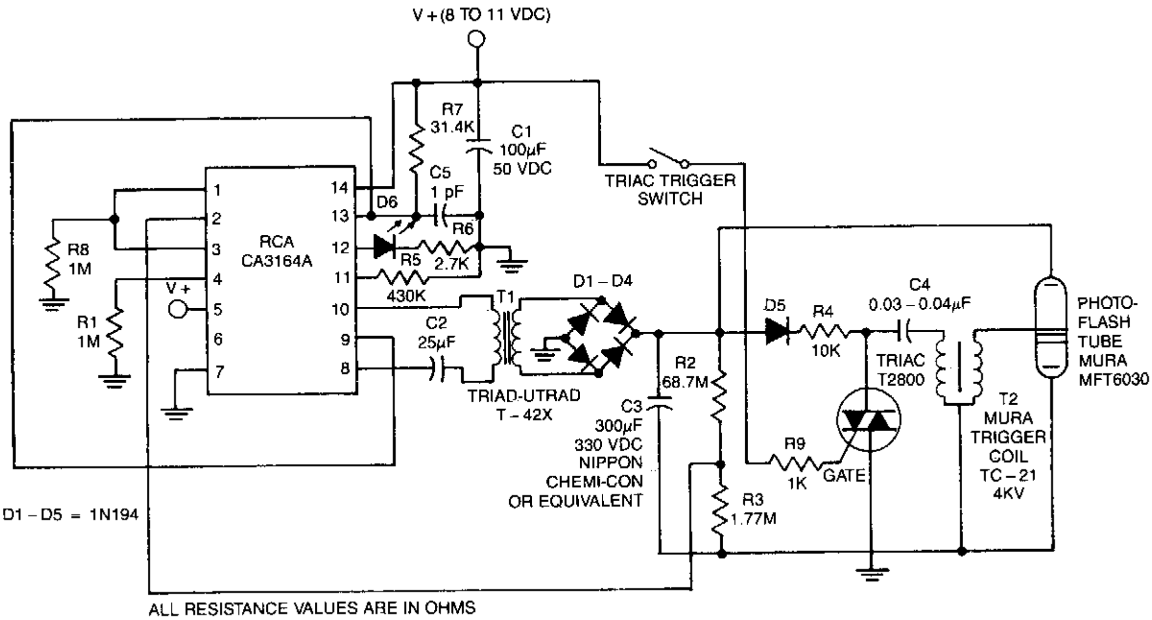
Power for the circuit is provided by a 7805 regulator. The unregulated 12 V output of BR1 is used to power the relay. The 12-V relay needs to have two sets of contacts as shown: to advance the projector, and to supply the beeps when recording. The LED indicates projector advance.

To record the beeps, connect beeper jack J2 to the input of the tape recorder and connect the controller to the projector-advance plug. The 60-Hz line frequency is used to produce beeps that are recorded on half of the stereo tape. The other track is used for commentary. The beeper output is controlled via 500-KΩ potentiometer R6.

Use pushbutton switch S1 to put the beeps on the tape, where required, to advance the projector. The beep length is automatic, and the projector will advance once for every push of S1.

When presenting your program, disconnect one speaker from the recorder, and connect the recorder to the jack on the controller and plug into the earphone jack. Connect the controller to the projector. The beeps will not be heard and the projector will advance at precisely the correct time.

ELECTRONIC PHOTOFLASH



GE/RCA

Fig. 71-7

Using the CA3164A BiMOS control chip consumes less than 15 μA during standby, yet it can provide 100 mA of chopped current to the dc-to-dc converter during the energy-reservoir charging cycle. The CA3164A drives the primary of T1 with symmetrically chopped current at a 400 to 2000 Hz rate. The CA3164A's chopper frequency is about 500 Hz; the duty cycle, 50% when R7 is 3.1 4 KΩ.

72

Power Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

50-W AUDIO POWER AMPLIFIER

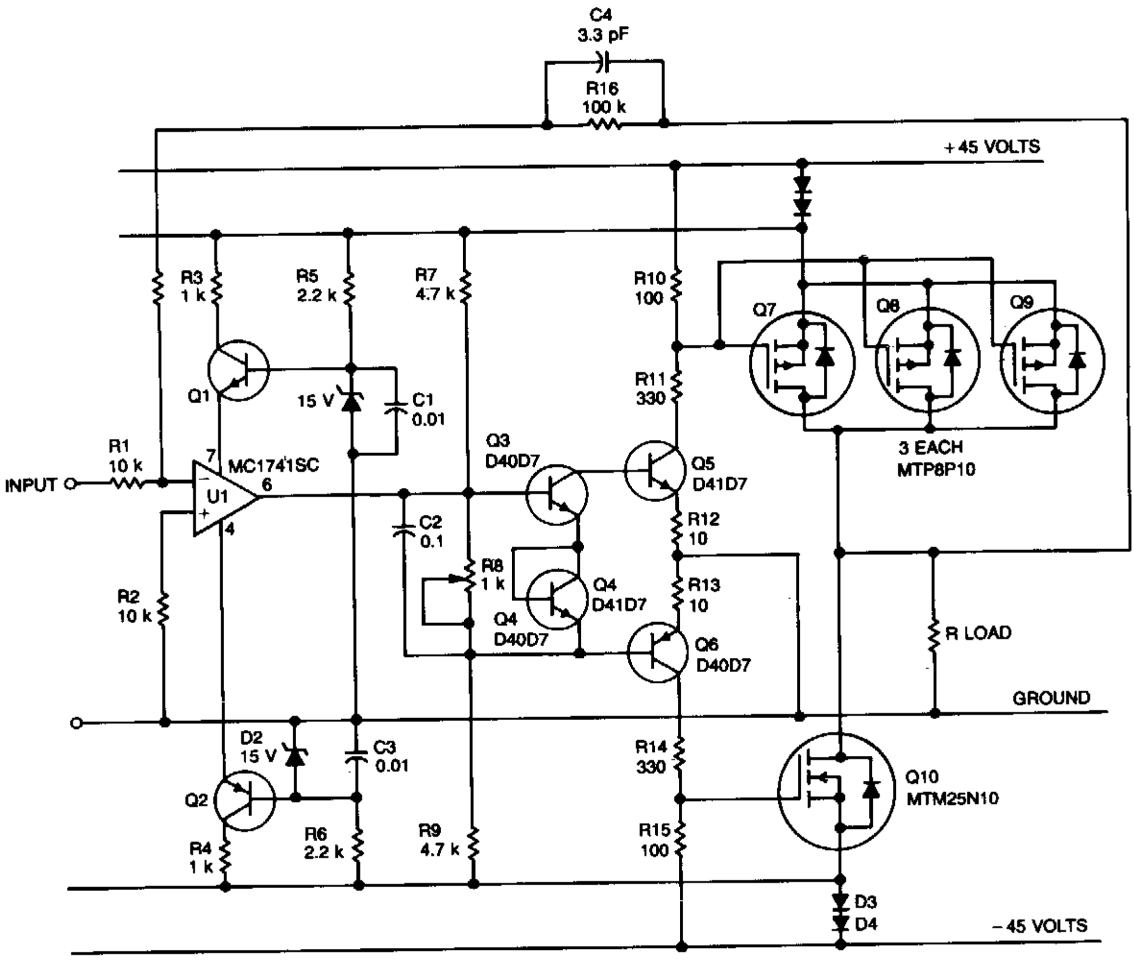
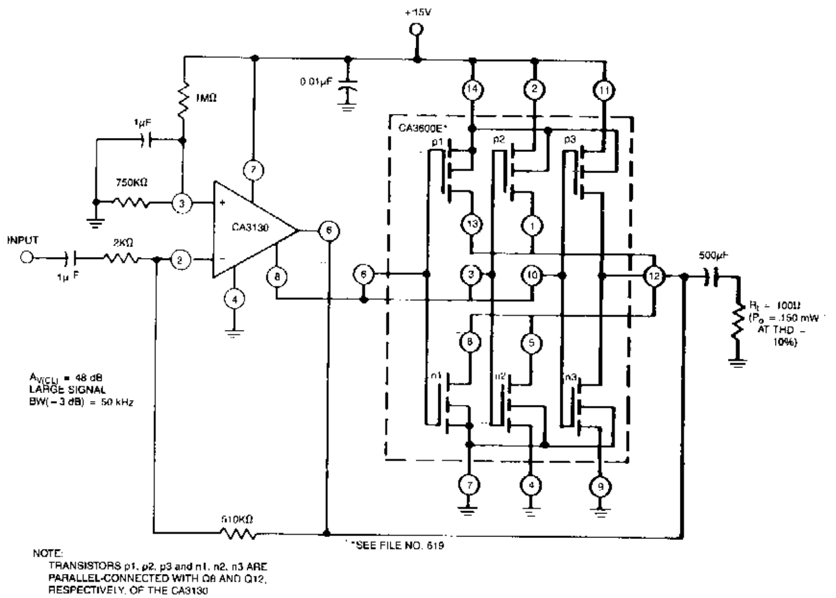


Fig. 72-1

This audio amplifier design approach employs TMOS Power FETs operating in a complementary common-source configuration. They are biased to cutoff, then turn on very quickly when a signal is applied. The advantage of this approach is that the output stage is very stable from a thermal point of view.

U1 is a high slew-rate amp that drives Q3, Q4,and Q6 (operating class AB) providing level transition for the output stage consisting of Q7, Q8, Q9,and Q10. The positive temperature coefficient of the TMOS device enables parallel operation of Q7, Q8, and Q9 and provides a higher power *complementary* device for Q10. These TMOS Power FETs must be driven from a low-source impedance of 100 Ω, in order to actually obtain high turn-on speeds.

OUTPUT-STAGE POWER BOOSTER



GE/RCA *Fig. 72-2*

This circuit easily supplements the current-sourcing and current-sinking capability of the CA3130 BiMOS op amp. This arrangement boosts the current-handling capability of the CA3130 output stage by about 2.5 times.

PORTABLE AMPLIFIER

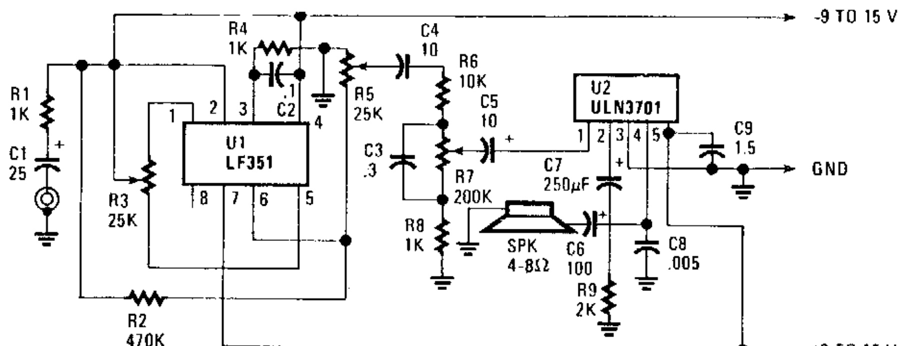
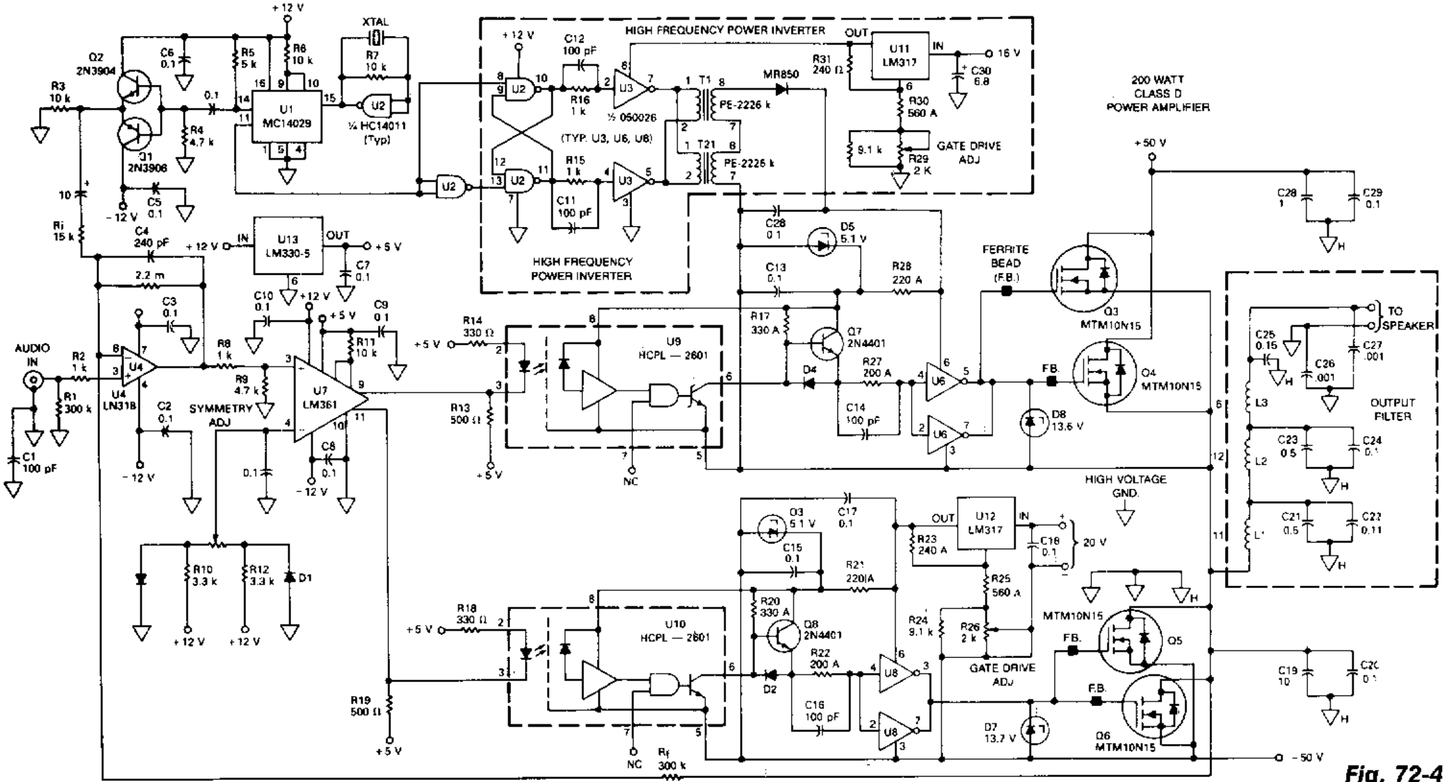


Fig. 72-3

POPULAR ELECTRONICS

U1, an FET op amp needs a bipolar voltage at pins 4 and 7 with a common ground for optimum gain. You can calculate the gain by dividing R2 by R1. Zero-set balance can be had through pins 1 and 5 through R3. Put a voltmeter between pin 6 and ground and adjust R3 for zero voltage. Once you've established that, you can measure the ohmic resistance at each side of R3's center tap and replace the potentiometer with fixed resistors. R6, R7, R8, and C3 form a tone control that will give you added bass boost, if needed.

CLASS-D POWER AMPLIFIER



Copyright of Motorola, Inc. Used by permission.

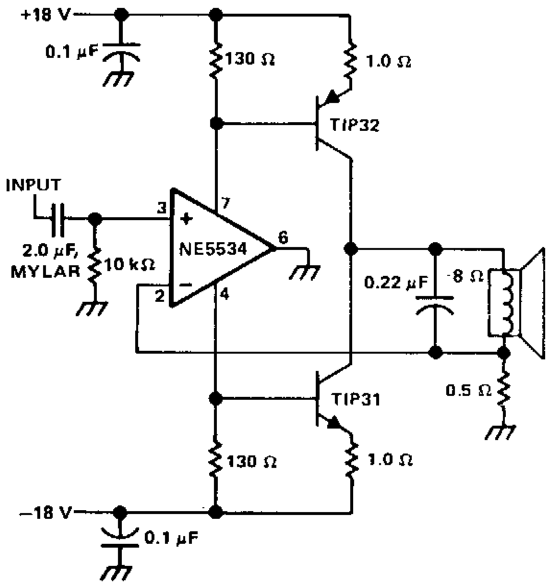
Fig. 72-4

In this circuit, a 2-MHz clock is divided by eight in U1, providing a stable 250-kHz carrier. Q1 and Q2 buffer the clock and provide a low-impedance drive for op amp U4, which is a high-gain amplifier and integrator. U4 accepts audio inputs and converts the 250-kHz square wave into a triangular wave. The summed audio and triangular-wave signal is applied to the input of comparator U7, where it is compared with a dc reference to produce a pulse-width modulated signal at the output of U7.

The output devices switch between the +50 V and −50 V rails in a complementary fashion, driving the output filter that is a sixth-order Butterworth low-pass type, which demodulates the audio and attenuates the carrier and high frequency components. Feedback is provided R_f; amplifier gain is $R_f R_i$.

Specifications: 200 W continuous power into a 4-Ω load; 20 to 20 kHz frequency response +0.5, −1.0 dB at 200 W; THD, IMD 0.5% at 200 W; 1.5-V rms input for rated output; 69 dB S/N ratio, A weighting; 6.6-V ms slew rate.

AUDIO POWER AMPLIFIER

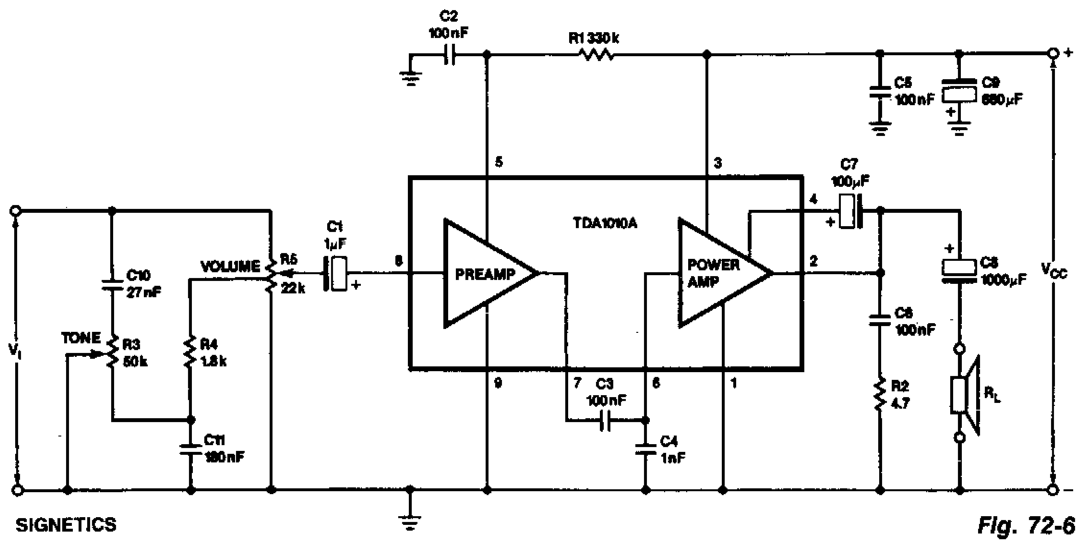


Reprinted by permission of Texas Instruments.

Fig. 72-5

The single speaker amplifier circuit uses current feedback, rather than the more popular voltage feedback. The feedback loop is from the junction of the speaker terminal and a 0.5-Ω resistor, to the inverting input of the NE5534. When the input to the amplifier is positive, the power supply supplies current through the TIP32 and the load to ground. Conversely, with a negative input, the TIP31 supplies current through the load to ground. The gain is set to about 15 (gain = SPKR 8 Ω/0.5 Ω feedback). The 0.22-μF capacitor across the speaker rolls off its response beyond the frequencies of interest. Using the 0.22-μF capacitor specified, the amplifier current output is 3 dB down at 90 kHz where the speaker impedance is about 20 Ω. To set the recommended class A output collector current, adjust the value of either 130-Ω resistor. An output current of 50 to 100 mA will provide a good operating midpoint between the best crossover distortion and power dissipation.

6-W AUDIO AMPLIFIER WITH PREAMP

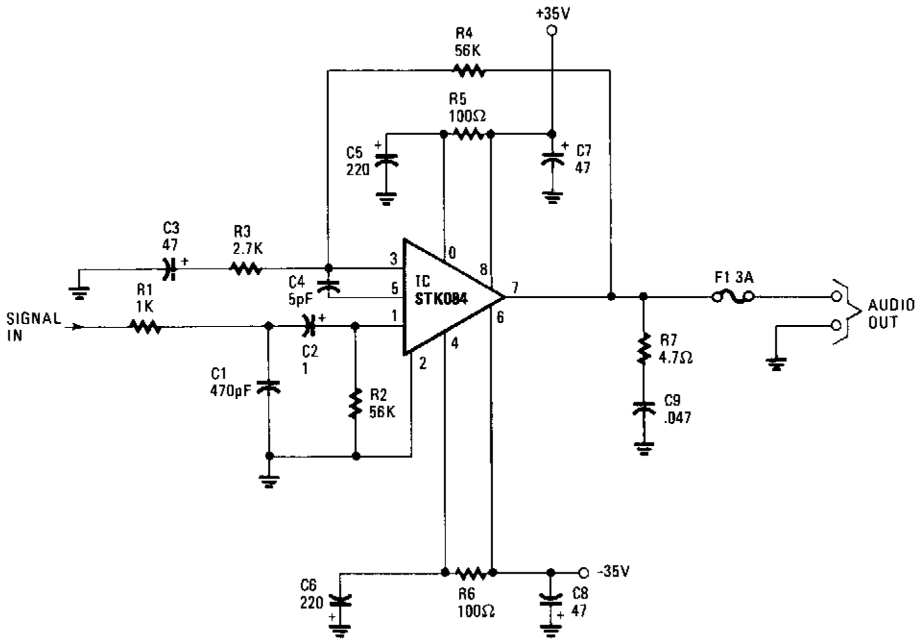


SIGNETICS

Fig. 72-6

This monolithic IC, class-B, audio amplifier circuit is a 6-W car radio amplifier for use with 4-Ω and 2-Ω load impedances.

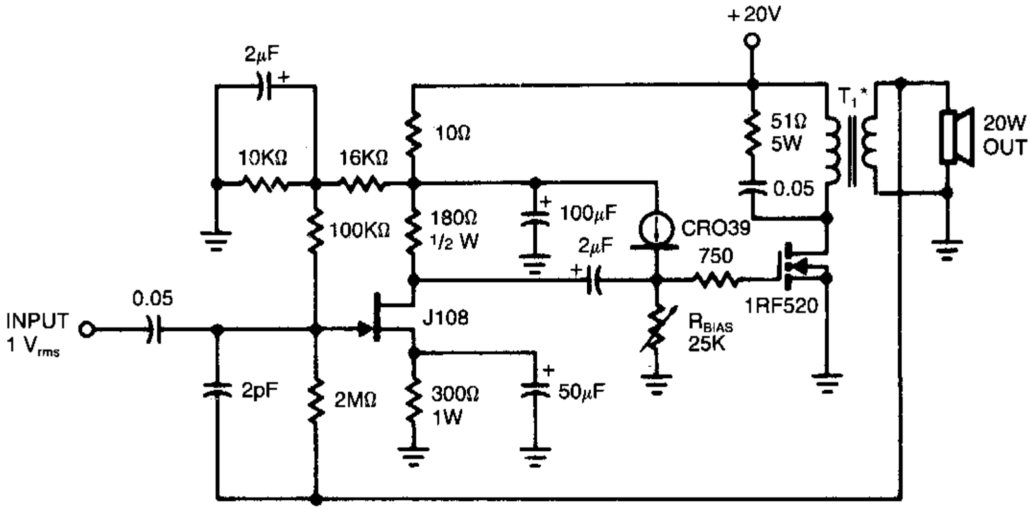
HYBRID POWER AMPLIFIER



HANDS-ON ELECTRONICS *Fig. 72-7*

The input is ac coupled to the amplifier through C2, which blocks dc signals that might also be present at the input. The R1/C1 combination forms a low-pass filter, which eliminates unwanted high-frequency signals by bypassing them to ground when they appear at the circuit input, which has an impedance of about 52 Ω. The gain of the amplifier is set at about 26 dB by resistors R3 and R4. The R5/C5/C7 combination on the positive supply and its counterpart R6/C6/C8 on the negative supply provides power-supply decoupling. R7 and C9 together prevent oscillation at the output of the amplifier. From that point, the amplifier's output signal is direct coupled to the speaker through a 3-A fuse, F1. The dc output of the amplifier at pin 7 is 0 V, so no dc current flows through the speaker. Should there be a catastrophic failure of the output stage, fuse F1, which should be a fast-acting type, prevents dc from flowing through the speaker.

20-W AUDIO AMPLIFIER



ALL RESISTORS 1/4 WATT, UNLESS NOTED
*TRIAD TY67A

SILICONIX ***Fig. 72-8***

This amplifier delivers 20 W into an 8-Ω load using a single IRF520 driving a transformer coupled output stage. This circuit is similar to the audio output stage used in many inexpensive radios and phonographs. Distortion is less than 5% at 10 W, using very little feedback (3%), with the IRF520 biased at 3 A.

73

Fixed Power Supplies

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SWITCHING POWER SUPPLY

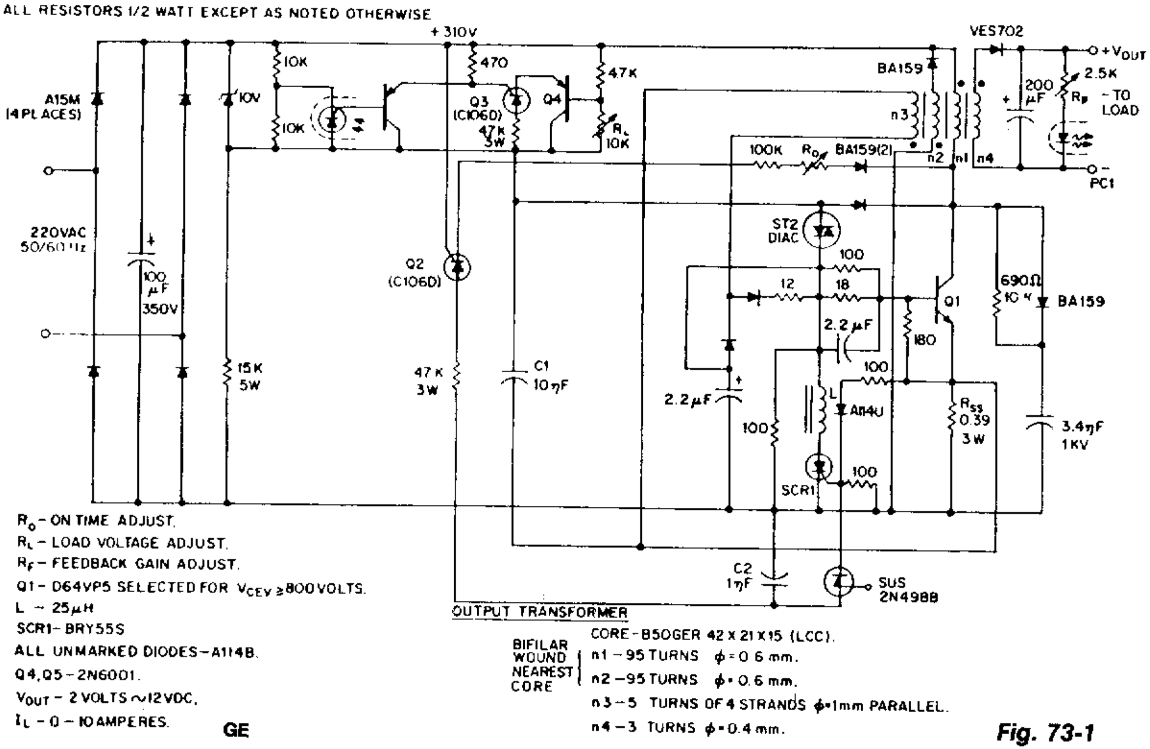


Fig. 73-1

This low-voltage high-current output, switching dc power supply is running off the 220-Vac input. In this circuit, an ST2 diac relaxation oscillator, Q3, C1, and the diac, initiates conduction of the output switching transistor Q1, the on-time of which is maintained constant by a separate timing/commutation network consisting of Q2, C2, SUS, and SCR 1. The output voltage, consequently, is dependent on the duty cycle. To compensate for unwanted variations of output voltage because of input voltage or load resistance fluctuations, an H11C wired as a linear-model unilateral PNP transistor in a stable differential amplifier configuration is connected into the galvanically isolated negative-feedback loop. The loop determines the duty cycle and hence the output voltage. Of further interest in this circuit is the use of several low-current, high-voltage, 400V V_{DRM} thyristors (Q2, Q3,) which are also used as pnp remote-base transistors. Short-circuit protection is assured by coupling Q1 collector-current feedback into the turn-off circuitry via R_{SS}.

LOW COST, LOW DROPOUT LINEAR REGULATOR

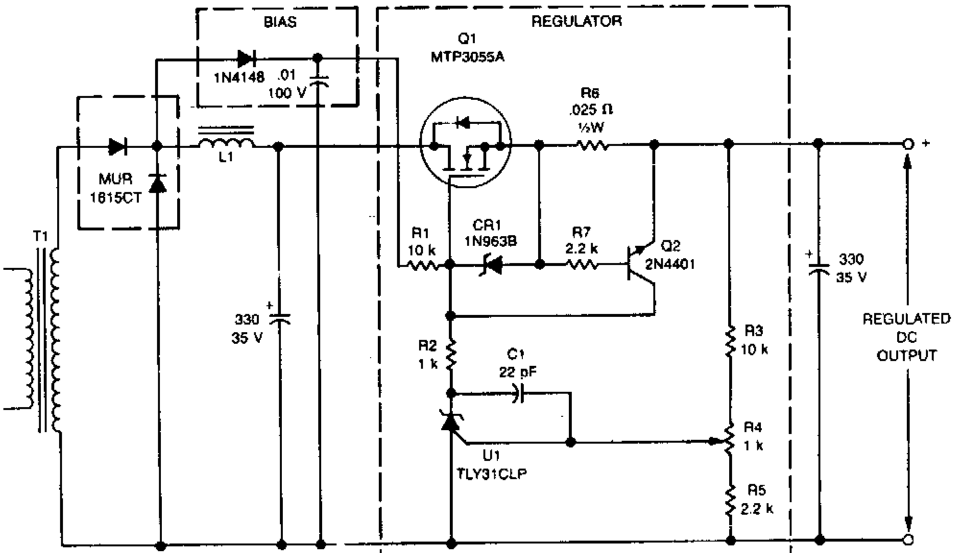


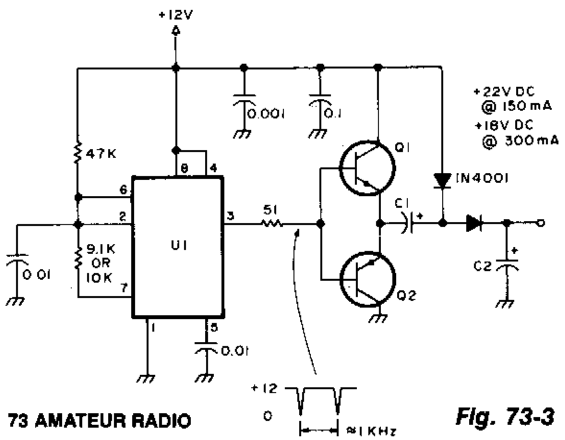
Fig. 73-2

This linear post regulator provides 12 V at 3 A. It employs TL431 reference U1 which, without additional amplification, drives TMOS MTP3055A gate Q1 series pass regulator. Bias voltage is applied through R1 to Q1's gate, which is protected against overvoltage by diode CR1. Frequency compensation for closed-loop stability is provided by C1.

Key performance features are:

Dropout voltage:	0.6 V	Load regulation:	10 mV
Line regulation:	± 5 mV	Output ripple:	10 mV pk-pk

VOLTAGE DOUBLER



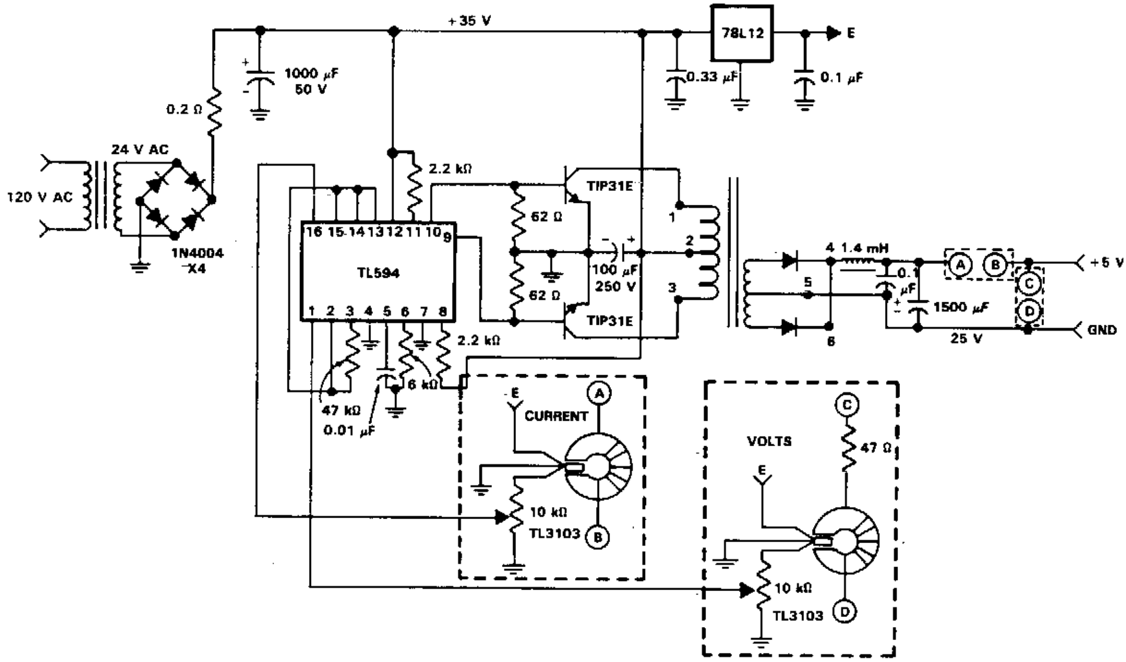
This circuit drives relays of 24 and 18 Vdc from a 12-V power supply. Use this circuit with almost any pnp or npn power transistor.

Parts: U1: NE 555 timer. C1 and C2: 50 μF/25 Vdc. Q1: TIP 29, TIP120, 2N4922, TIP61, TIP110, or 2N4921. Q2: TIP30, TIP125, 2N4919, TIP62, TIP115, or 2N4918.

73 AMATEUR RADIO

Fig. 73-3

ISOLATED FEEDBACK POWER SUPPLY

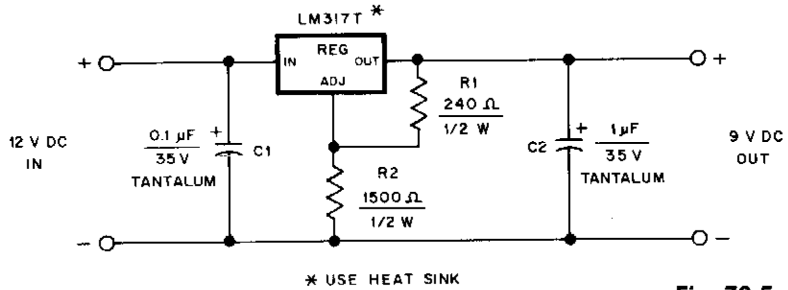


Reprinted by permission of Texas Instruments.

Fig. 73-4

Figure 73-4 is a power supply circuit using the isolated feedback capabilities of the TL3103 for both current and voltage sensing. This supply is powered from the ac power line and has an output of 5 V at 1.5 A. Both output voltage and current are sensed and the error voltages are applied to the error amplifiers of the TL594 PWM control IC. The 24-V transformer produces about 35 V at the 1000-μF filter capacitor. The 20-kHz switching frequency is set by the 6-KΩ resistor and the 0.01-μF capacitor on pins 6 and 5, respectively. The TL594 is set for push-pull operation by typing pin 13 high. The 5-V reference on pin 14 is tied to pin 15, which is the reference or the current error amplifier. The 5-V reference is also tied to pin 2 which is the reference for the output voltage error amplifier. The output voltage and current limit are set by adjustment of the 10-KΩ pots in the TL3103 error-sensing circuits. A pair of TIP31E npn transistors are used as switching transistors in a push-pull circuit.

HAND-HELD TRANSCEIVER DC ADAPTER

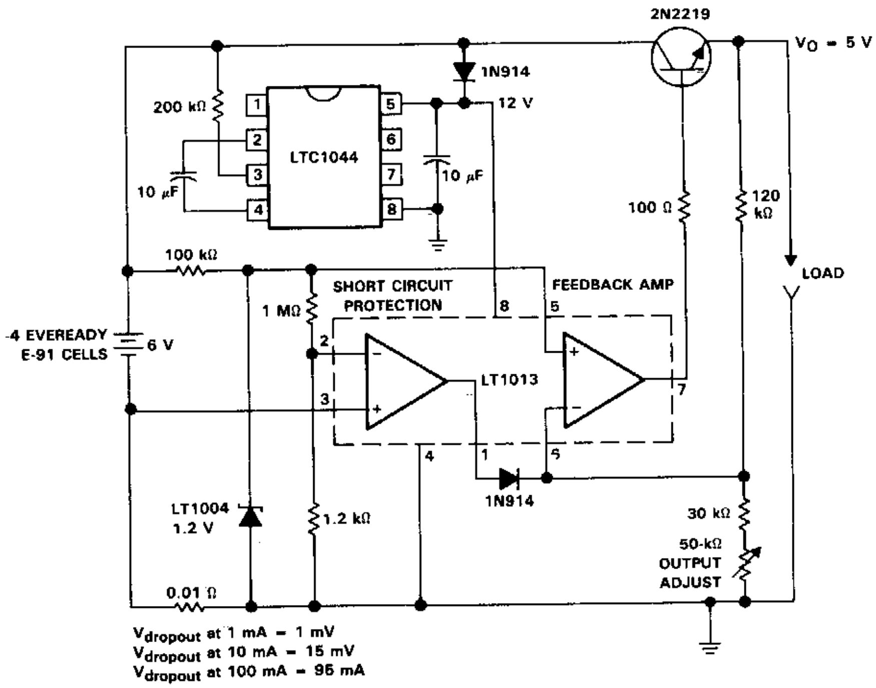


QST

Fig. 73-5

This dc adapter provides a regulated 9-V source for operating a Kenwood TR-2500 hand-held transceiver in the car. The LM317T's mounting tab is electrically connected to its output pin, so take this into account as you construct your version of the adapter. The LM317T regulator dissipates 2 or 3 W in this application, so mount it on a 1-×-2-inch piece of 1/8-inch-thick aluminum heatsink.

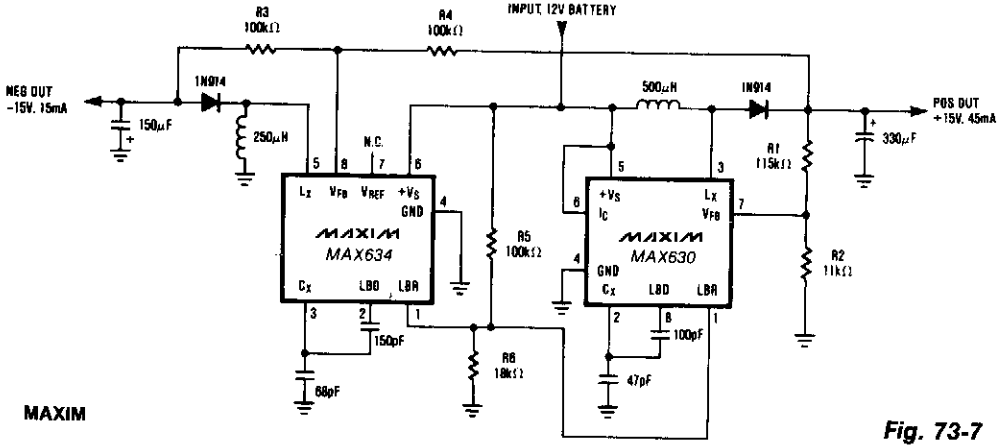
LOW-DROPOUT 5 V REGULATOR



Reprinted by permission of Texas Instruments.

Fig. 73-6

DUAL-TRACKING REGULATOR

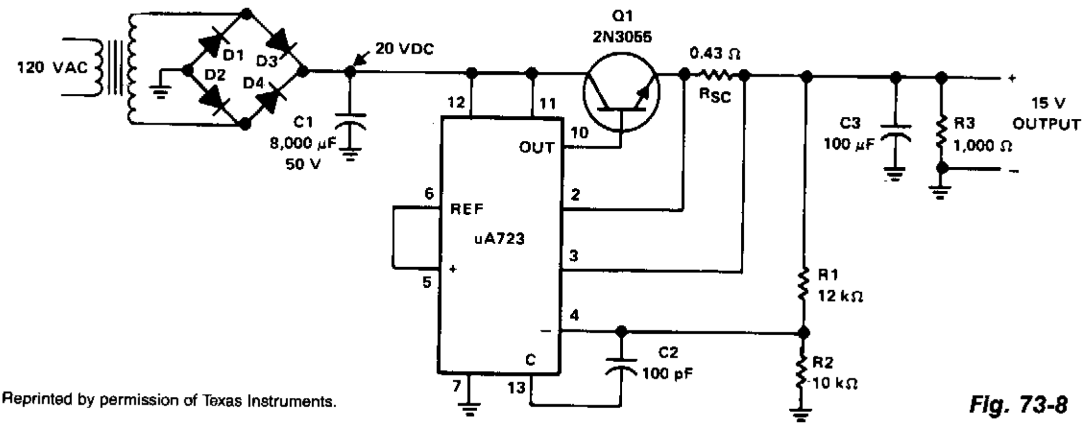


MAXIM

Fig. 73-7

A MAX634 inverting regulator is combined with a MAX630 to provide a dual tracking ±15 output from a 12-V battery. The reference for the −15 V output is derived from the positive output via R3 and R4. Both regulators are set to maximize output power at low battery voltages by reducing the oscillator frequency, via LBR, when V_{BATT} falls to 8.5 V.

+15 V 1-A REGULATED POWER SUPPLY



Reprinted by permission of Texas Instruments.

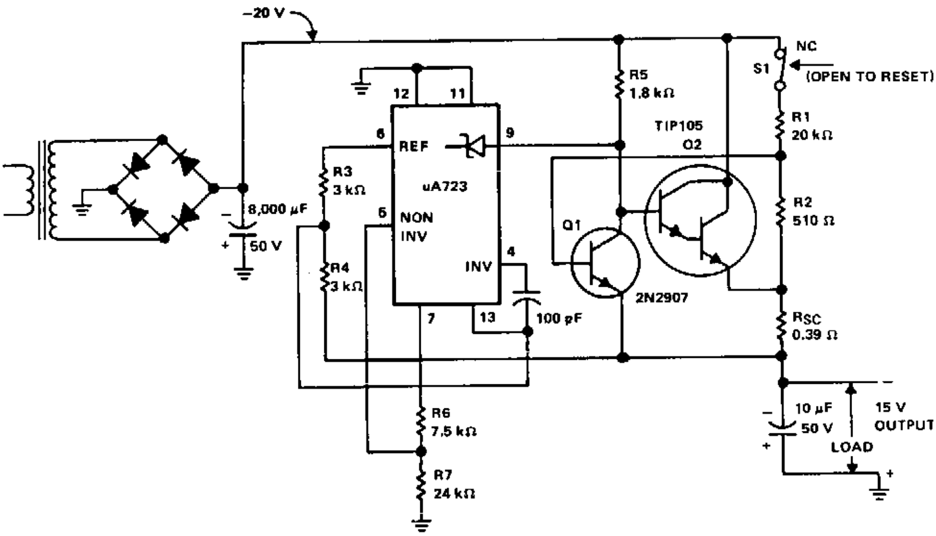
Fig. 73-8

The supply receives +20 Vdc from the rectifier/filter section. This is applied to pins 11 and 12 of the uA723, as well as to the collector of the 2N3055 series-pass transistor. The output voltage is sampled through R1 and R2, providing about 7 V with respect to ground at pin 4. The reference terminal at pin 6 is tied directly to pin 5, the noninverting input of the error amplifier. For fine trimming the output voltage, a potentiometer can be installed between R1 and R2. A 100-pF capacitor from pin 13 to pin 4 furnishes gain compensation for the amplifier.

Base drive to the 2N3055 pass transistor is furnished by pin 10 of the uA723. Since the desired output of the supply is 1 A, maximum current limit is set to 1.5 A by resistor R_{SC} whose value is 0.433 Ω.

A 100-μF electrolytic capacitor is used for ripple voltage reduction at the output. A 1-KΩ output resistor provides stability for the power supply under no-load conditions. The 2N3055 pass transistor must be mounted on an adequate heatsink.

−15 V 1-A REGULATED POWER SUPPLY

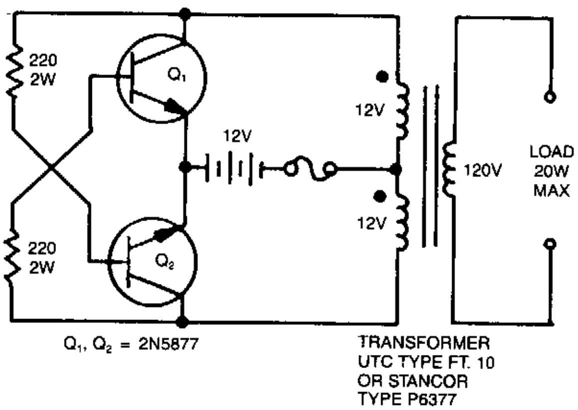


Reprinted by permission of Texas Instruments.

Fig. 73-9

The supply receives −20 V from the rectifier/filter which is fed to the collector of the Darlington pnp pass transistor, a TIP105. The base drive to the TIP105 is supplied through resistor R5. The base of the TIP105 is driven from V_Z terminal at pin 9, which is the anode of a 6.2-V zener diode that connects to the emitter of the uA723 output control transistor. The method of providing the positive feedback required for foldback action is shown. This technique introduces positive feedback by increased current flow through resistors R1 and R2 under short-circuit conditions. This forward biases the base-emitter junction of the 2N2907 sensing transistor, which reduces base drive to the TIP105.

12-VDC BATTERY-OPERATED 120-VAC POWER SOURCE



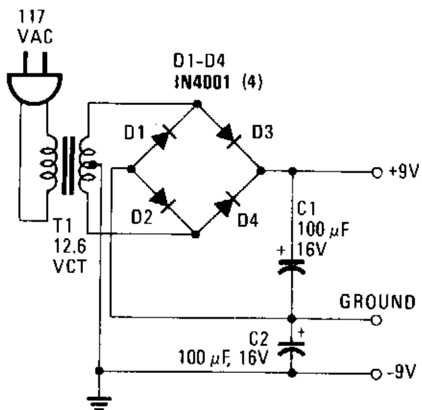
EDN

Fig. 73-10

A simple 120 V: 24 V, center-tapped control transformer and four additional components can do the job. This circuit outputs a clean 200 V pk-pk square wave at 60 Hz and can supply up to 20 W. The circuit is self-starting and free-running.

If Q1 is faster and has a higher gain than Q2, it will turn on first when you apply the input power and will hold Q2 off. Load current and transformer magnetizing current then flows in the upper half of the primary winding, and auto transformer action supplies the base drive until the transformer saturates. When that action occurs, Q1 loses its base drive. As it turns off, the transformer voltages reverse, turning Q2 on and repeating the cycle. The output frequency depends on the transformer iron and input voltage, but not on the load. The frequency will generally range between 50 to 60 Hz with a 60-Hz transformer and car battery or equivalent source. The output voltage depends on turns ratio and the difference between input voltage and transistor saturation voltage. For higher power, use larger transformers and transistors. This type of inverter normally is used in radios, phonographs, hand tools, shavers, and small fluorescent lamps. It will not work with reactive loads (motors) or loads with high inrush currents, such as coffee pots, frying pans, and heaters.

SIMPLE POWER SUPPLY



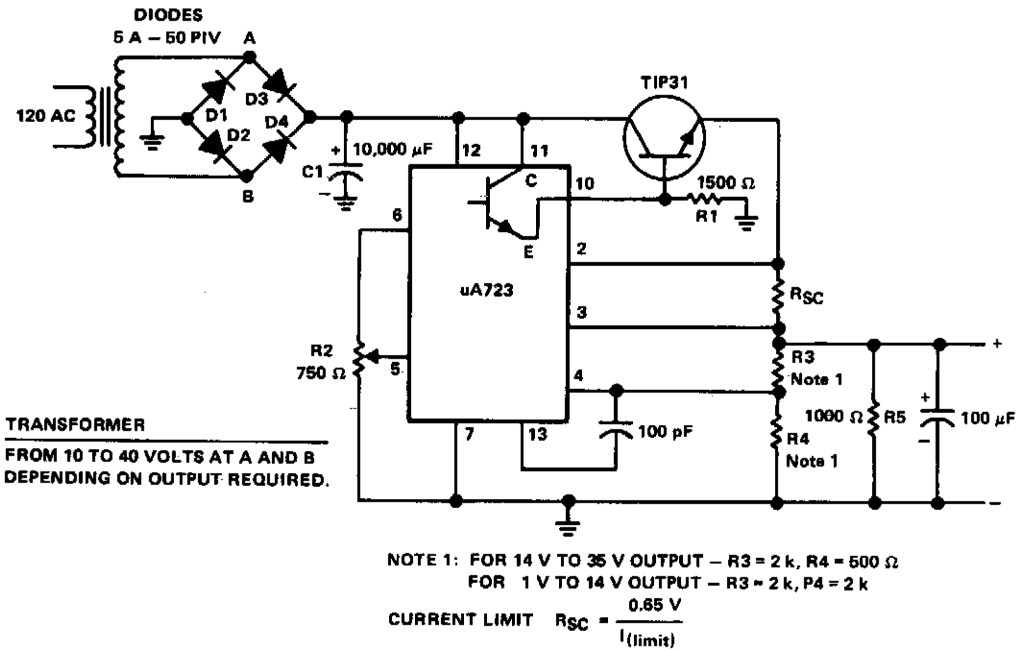
HANDS-ON ELECTRONICS

Fig. 73-11

This power supply delivers plus and minus 9 V to replace two 9-V batteries. The rectifier circuit is actually two separate full-wave rectifiers fed from the secondary of the transformer. One full-wave rectifier is composed of diodes D1 and D2, which develop +9 V, and the other is composed of D3 and D4, which develop −9 V.

Each diode from every pair rectifies 6.3 Vac, half the secondary voltage, and charges the associated filter capacitor to the peak value of the ac waveform, $6.3 \times 1.414 = 8.9$ V. Each diode should have a PIV, Peak Inverse Voltage, rating that is at least twice the peak voltage from the transformer, $2 \times 8.9 = 18$ V. The 1N4001 has a PIV of 50 V.

GENERAL-PURPOSE POWER SUPPLY



Reprinted by permission of Texas Instruments.

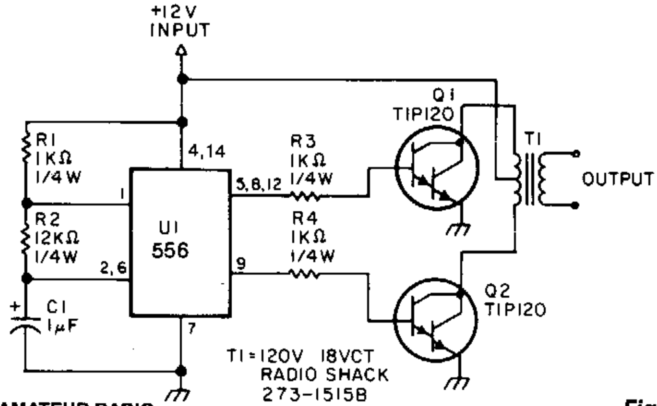
Fig. 73-12

The supply 6-66 can be used for supply output voltages from 1 to 35 V. The line transformer should be selected to give about 1.4 times the desired output voltage from the positive side of filter capacitor C1 to ground. Potentiometer R2 sets the output voltage to the desired value by adjusting the reference input. R_{SC} is the current limit set resistor. Its value is calculated as:

$$R_{SC} = \frac{0.65\ \text{V}}{I_L}$$

For example, if the maximum current output is to be 1 A, $R_{SC} = 0.65/1.0 = 0.65\ \Omega$. The 1-KΩ resistor, R_S, is a light-loaded resistor designed to improve the no-load stability of the supply.

LOW-POWER INVERTER

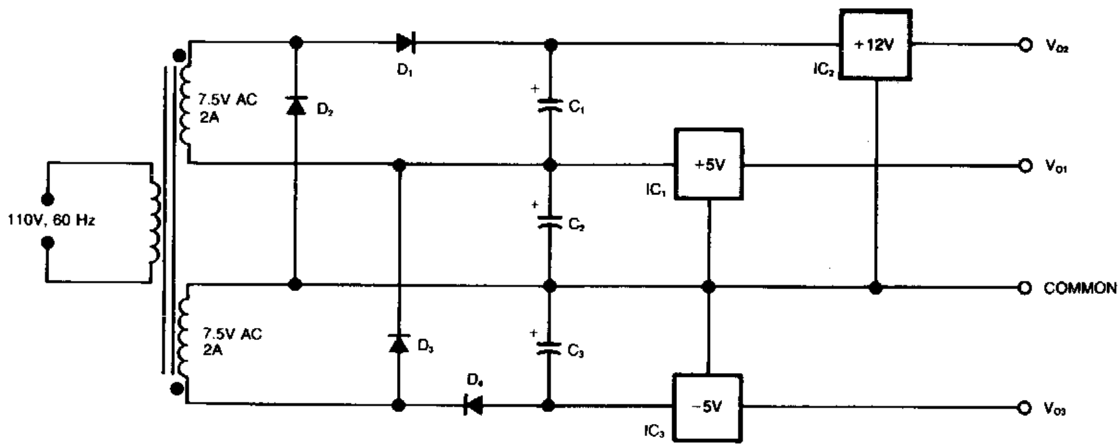


73 AMATEUR RADIO

Fig. 73-13

This low-power inverter uses only 9 parts and turns 10 to 16 Vdc into 60-Hz, 115-V square-wave power to operate ac equipment up to 25 W. The first section of the 556 timer chip is wired as an astable oscillator with R2 and C1 setting the frequency. The output is available at pin 5. The second section is wired as a phase inverter. That output is available at pin 9. Resistors R3 and R4 keep output transistors Q1 and Q2 from loading down the oscillator. The two transistors drive the transformer push-pull fashion. When one transistor is biased-on, the other is cut-off. The transformer is a 120 V/18 VCT unit that is connected backwards, so that it steps the voltage up rather than down. Oscillator circuit U1, R1, R2, and C1 operates from about 4 to 16 V with a very stable output.

THREE-RAIL POWER SUPPLY

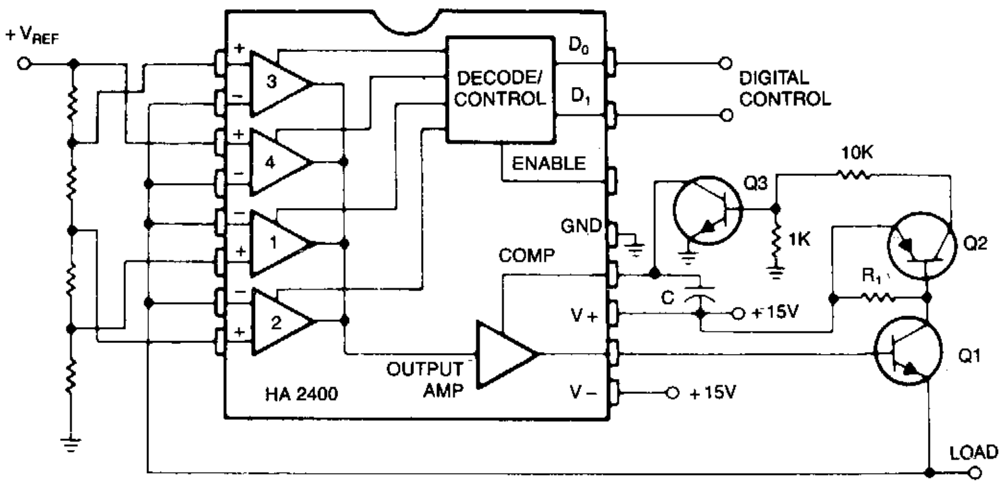


EDN

Fig. 73-14

This circuit generates three supply voltages using a minimum of components. Diodes D2 and D3 perform full-wave rectification, alternately charging capacitor C2 on both halves of the ac cycle. On the other hand, diode D1 with capacitor C1, and diode D4 with capacitor C3 each perform half-wave rectification. The full- and half-wave rectification arrangement is satisfactory for modest supply currents drawn from −5 and +12-V regulators IC3 and IC2. You can use this circuit as an auxiliary supply in an up-based instrument, for example, and avoid the less attractive alternatives of buying a custom-wound transformer, building a more complex supply, or using a secondary winding, say 18 Vac, and wasting power in the 5-V regulators.

PROGRAMMABLE POWER SUPPLY

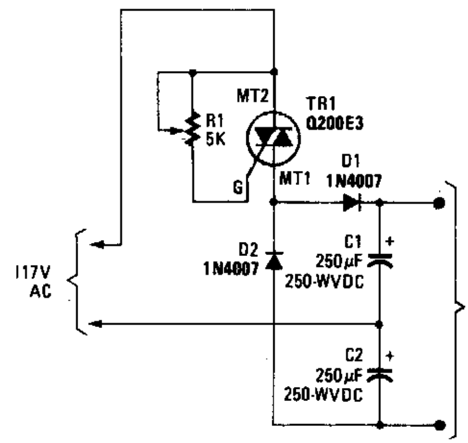


HARRIS

Fig. 73-15

Many systems require one or more relatively low-current voltage sources which can be programmed to a few predetermined levels. The circuit shown above produces positive output levels, but could be modified for negative or bipolar outputs. Q1 is the series regulator transistor, selected for the required current and power capability. R1, Q2, and Q3 form an optional short circuit protection circuit, with R1 chosen to drop about 0.7 V at the maximum output current. The compensation capacitor, C, should be chosen to keep the overshoot, when switching, to an acceptable level.

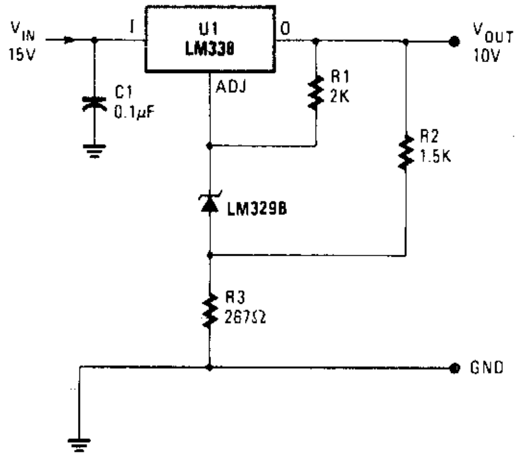
TRIAC-CONTROLLED VOLTAGE DOUBLER



HANDS-ON ELECTRONICS

Fig. 73-16

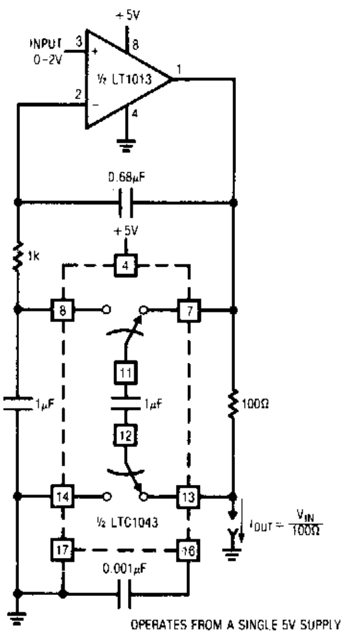
HIGH STABILITY 10-V REGULATOR



POPULAR ELECTRONICS

Fig. 73-17

VOLTAGE-CONTROLLED CURRENT SOURCE WITH GROUNDED SOURCE AND LOAD

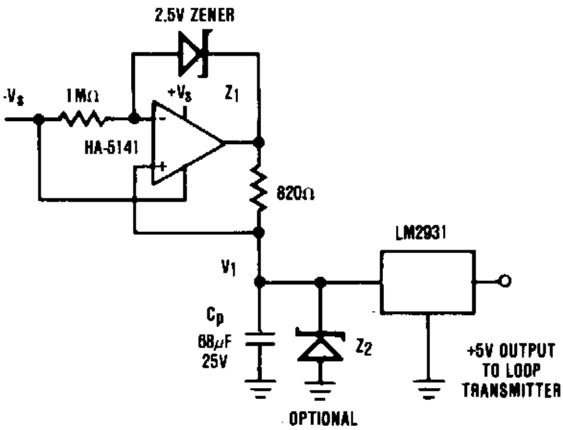


This is a simple, precise voltage-controlled current source. Bipolar supplies will permit bipolar output. Configurations featuring a grounded voltage-control source and a grounded load are usually more complex and depend upon several components for stability. In this circuit, accuracy and stability almost entirely depend upon the 100-Ω shunt.

LINEAR TECHNOLOGY CORP.

Fig. 73-18

CHARGE POOL POWER SUPPLY



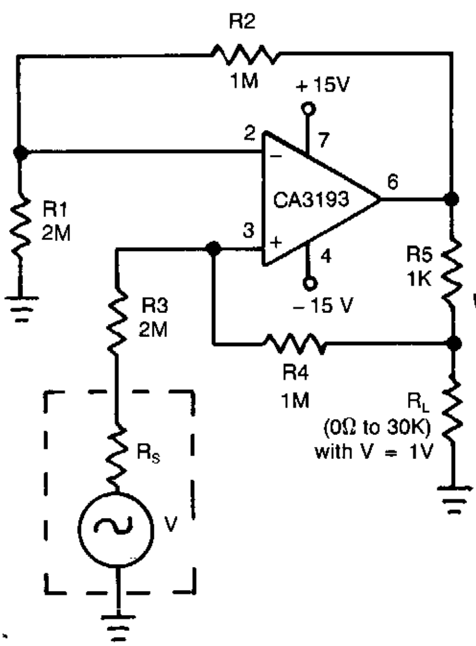
HARRIS

Fig. 73-19

It is usually desirable to have the remote transmitter of a 4 to 20 mA current loop system powered directly from the transmission line. In some cases, this is not possible because of the high-power requirements set by the remote sensor/transmitter system. In these cases, an alternative to the separate power supply is still possible. If the remote transmitter can be operated in a pulsed mode where it is active only long enough to perform its function, then a charge pool power supply can still allow the transmitter to be powered directly by the current loop. In this circuit, constant current *I*1 is supplied to the charge pool capacitor, CP, by the HA-5141 (where *I*1 = 3 mA). The voltage *V*1 continues to rise until the output of the HA-5141 approaches $+V_s$ or the optional voltage limiting provided by *Z*2. The LM2931 voltage regulator supplies the transmitter with a stable +5 V supply from the charge collected by CP. Available power supply current is determined by the duration, allowable voltage droop on CP, and required repetition rate. For example, if *V*1 is allowed to droop 4.4 V and the duration of operation is 1 ms, the available power supply current is approximately:

$$= CP\frac{dV_1}{dt} = 68\mu\text{F} \times \frac{4.4\text{ V}}{1\ ms} = 30\ mA$$

BILATERAL CURRENT SOURCE



ALL RESISTORS ARE 1%
ALL RESISTANCE VALUES ARE IN OHMS
IF R1 = R3 AND R2 ≈ R4 + R5 THEN

I_L IS INDEPENDENT OF VARIATIONS IN R_L
FOR R_L VALUES OF 0Ω to 3KΩ WITH V = 1V

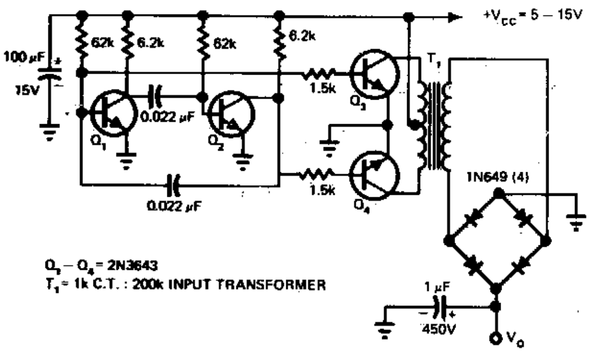
$$I_L = \frac{V\ R4}{R3\ R5} = \frac{V\ 1M}{(2M)(1K)} = \frac{V}{2K} = 500\mu A$$

GE/RCA

Fig. 73-20

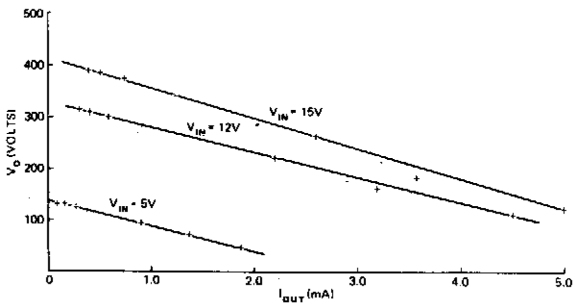
This circuit uses a CA3193 precision op amp to deliver a current independent of variations in R_L. With *R*1 set equal to *R*3, and *R*2 approximately equal to *R*4 + *R*5, the output current, I_L, is: V_{IN} (*R*4)/(*R*3) (*R*5). 500-μA load current is constant for load values from 0 to 3 Ω.

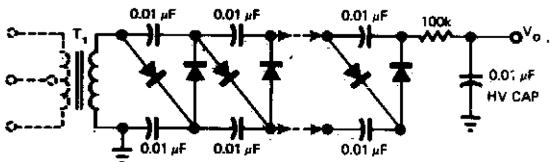
POWER CONVERTER



INPUT AND OUTPUT PARAMETERS FOR LOW COST POWER CONVERTER

R_L (OHMS)	V_O (VOLTS)	I_O (mA)	I_{IN} (mA)	P_O (W)	P_{IN} (W)	EFF. (%)
V_{IN} = 5V						
0	160	0	22	0	0.11	–
1 M	130	0.13	25	0.017	0.125	14
750k	130	0.17	26	0.022	0.130	17
510k	125	0.25	27	0.031	0.135	23
100k	90	0.90	42	0.081	0.210	39
51k	70	1.37	60	0.096	0.300	32
24k	45	1.88	60	0.085	0.300	38
V_{IN} = 12V						
0	415	0	60	0	0.720	–
1 M	315	0.32	65	0.10	0.78	13
750k	310	0.41	65	0.13	0.78	17
510k	300	0.59	70	0.18	0.84	21
100k	220	2.20	100	0.48	1.20	40
51k	165	3.24	120	0.54	1.44	38
24k	110	4.58	140	0.50	1.68	30
V_{IN} = 15V						
0	520	0	80	0	1.2	...
1 M	390	0.39	80	0.15	1.2	13
750k	385	0.51	85	0.196	1.28	15
510k	375	0.74	90	0.28	1.35	21
100k	260	2.6	120	0.68	1.8	38
51k	180	3.57	140	0.65	2.1	31
24k	120	5.0	160	0.60	2.4	25





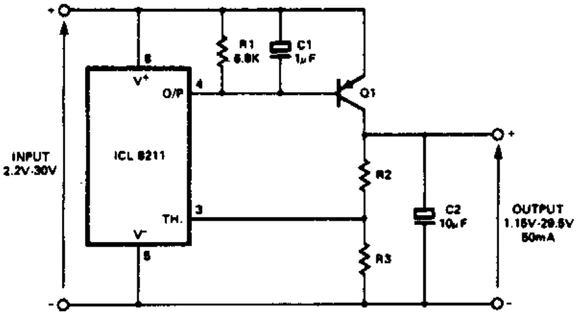
- ADD ADDITIONAL STAGES AS REQUIRED
- DIODES ARE 1N649 OR EQUIVALENT
- CAPACITORS ARE CERAMIC DISC 1 kV

EDN

Fig. 73-21

This circuit consists of an astable multivibrator driving a push-pull pair of transistors into the transformer primary. The multivibrator frequency should equal around 1 or 2 kHz. For higher dc voltages, voltage multipliers on the secondary circuit have been used successfully to generate 10 kV from a 40-stage multiplier like the one shown.

POSITIVE REGULATOR WITH PNP BOOST



INTERSIL **Fig. 73-22**

The IC8211 provides the voltage reference and regulator amplifier, while Q1 is the series pass transistor. R1 defines the output current of the IC8211, while C1 and C2 provide loop stability and also act to suppress feedthrough of input transients to the output supply. R2 and R3 determine the output voltage as follows:

output supply. R2 and R3 determine the output voltage as follows:

$$V_{\mathrm{OUT}} = 1.5 \times \frac{R_2 + R_3}{R_3}$$

In addition, the values of R2 and R3 are chosen to provide a small amount of standing current in Q1, which gives additional stability margin to the circuit. Where accurate setting of the output voltage is required, either R2 or R3 can be made adjustable. If R2 is made adjustable, the output voltage will vary linearly with the shaft angle; however, if the potentiometer wiper was to open the circuit, the output voltage would rise. In general, therefore, it is better to make R3 adjustable, since this gives fail-safe operation.

LOW FORWARD-DROP RECTIFIER CIRCUIT

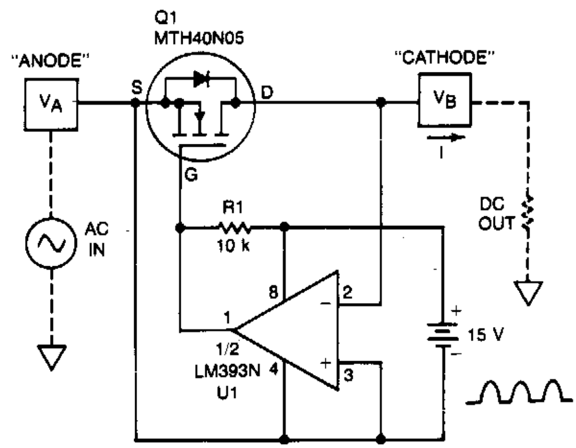
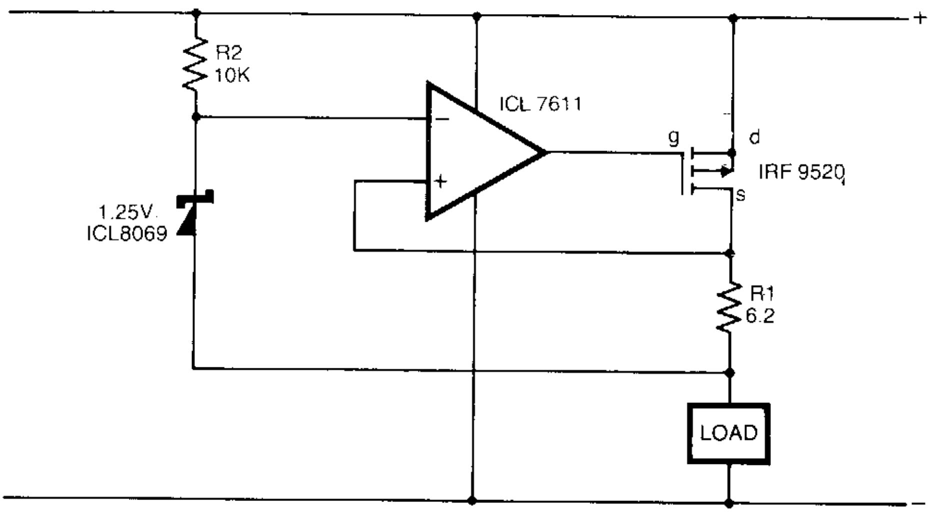


 Fig. 73-23

A TMOS power FET, Q1, and an LM393 comparator provide a high-efficiency rectifier circuit. When V_A exceeds V_B, U1's output becomes high and Q1 conducts. Conversely, when V_B exceeds V_A, the comparator output becomes low and Q1 does not conduct.

The forward drop is determined by Q1's on resistance and current I. The MTH40N05 has an on resistance of 0.028 Ω; for I = 10 A, the forward drop is less than 0.3 V. Typically, the best Schottky diodes do not even begin conducting below a few hundred mV.

SAFE CONSTANT-CURRENT SOURCE

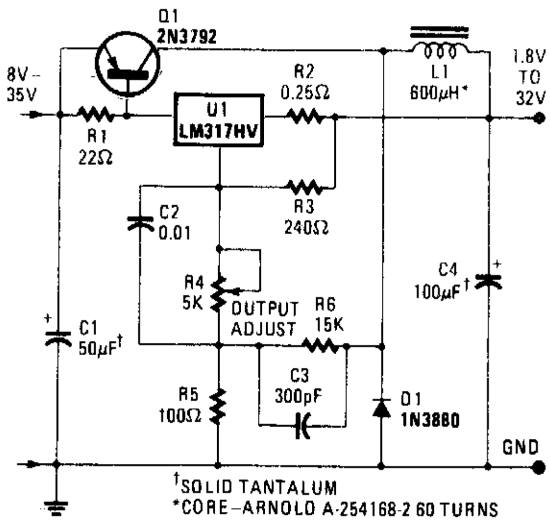


ELECTRONIC ENGINEERING

Fig. 73-24

In the circuit shown, a CMOS op amp controls the current through a p-channel HEXFET power transistor to maintain a constant voltage across R1. The current is given by: $1 = V_{REF}/R1$. The advantages of this configuration are: (a) in the event of a component failure, the load current is limited by R1; and (b) the overhead voltage needed by the op amp and the HEXFET is extremely low.

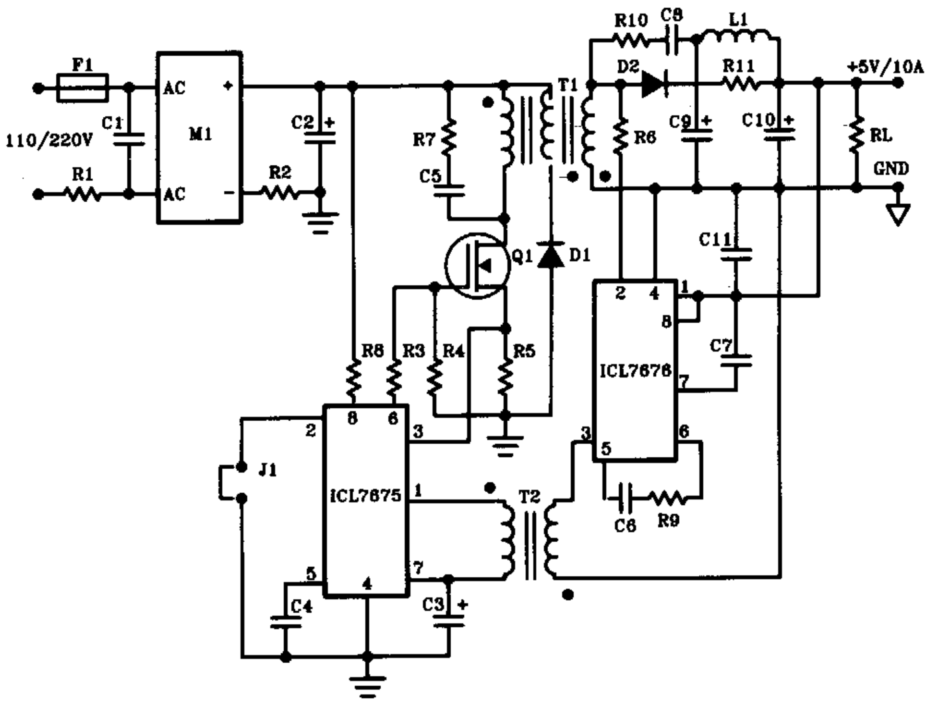
LOW-COST 3-A SWITCHING REGULATOR



POPULAR ELECTRONICS

Fig. 73-25

50-W OFF-LINE SWITCHING POWER SUPPLY



Component Values Table

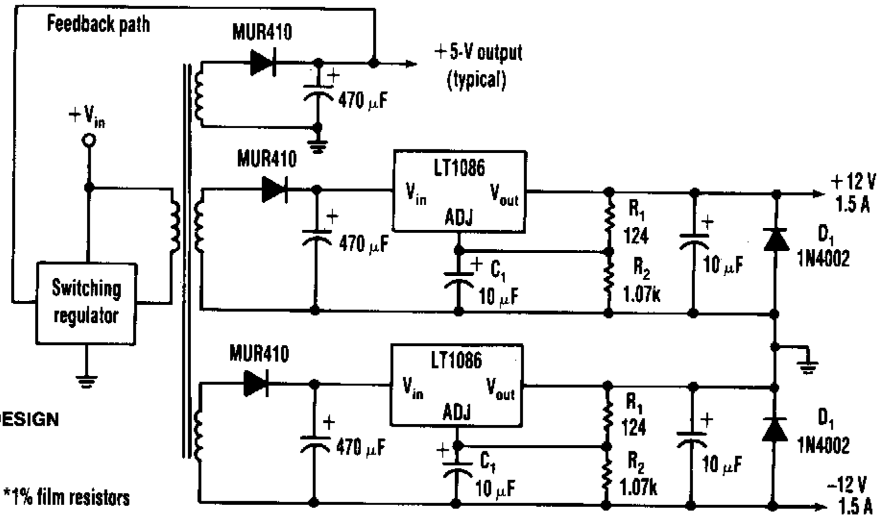
C1	0.022 μF/400V	R1	100Ω at 25°C	L1	25 μH
C2	470 μF/250V	R2	1Ω/1W	D1	1N4937
C3	470 μF/16V	R3	10Ω/0.25W	D2	MBR1035
C4	220 pF/100V	R4	100 kΩ/0.25W	T1	Lp=.9 mH, n=1:15
C5	470 pF/500V	R5	0.33Ω/1W	T2	50 μH, n=1:3
C6	2200 pF/500V	R6	10 kΩ/0.25W	F1	Fuse 1 A/SB
C7	270 pF/500V	R7	390Ω/2W	M1	Diode Bridge
C8	39 pF/500V	R8	22 kΩ/10W	Q1	BUZ80A/IXTP4N80 (220VAC)
C9	11,000 μF/6.3V	R9	68Ω/0.25W		
C10	10 μF/16V	R10	10Ω/0.5W	Q1	GE IRF823 (110VAC)
C11	0.047 μF/10V	R11	3.3Ω/0.5W		
		RL	5Ω/10W		

INTERSIL

Fig. 73-26

The schematic shows a 50-W power supply with a 5-V 10-A output. It is a flyback converter operating in the continuous mode. The circuit features a primary side and secondary side controller will full-protection from fault conditions such as overcurrent. After the fault condition has been removed, the power supply will enter the soft-start cycle before recommencing normal operation.

EFFICIENT NEGATIVE VOLTAGE REGULATOR



ELECTRONIC DESIGN

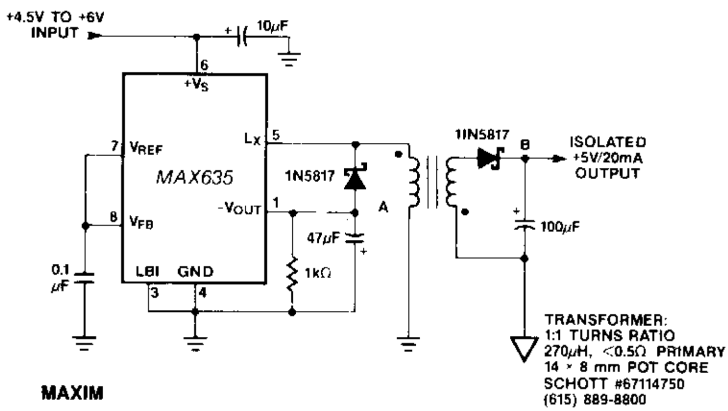
Fig. 73-27

One way to provide good negative-voltage regulation is with a low-dropout positive-voltage regulator operating from a well-isolated secondary winding of switch-mode circuit transformer. The technique works with any positive-voltage regulator, although highest efficiency occurs with low-dropout types.

Under all loading conditions, the minimum voltage difference between the regulator V_{IN} and V_{OUT} pins must be at least 1.5 V, the LT1086's low-dropout voltage. If this requirement isn't met, the output falls out of regulation. Two programming resistors, R1 and R2, set the output voltage to 12 V, and the LT1086's servo the voltage between the output and its adjusting (ADJ) terminals to 1.25 V. Capacitor C1 improves ripple rejection, and protection diode D1 eliminates common-load problems.

Since a secondary winding is galvanically isolated, a regulator's 12 V output can be referenced to ground. Therefore, in the case of a negative-voltage output, the positive-voltage terminal of the regulator connects to ground, and the −12 V output comes off the anode of D1. The V_{IN} terminal floats at 1.5 V or more above ground.

5 V-TO-ISOLATED 5 V AT 20 MA CONVERTER

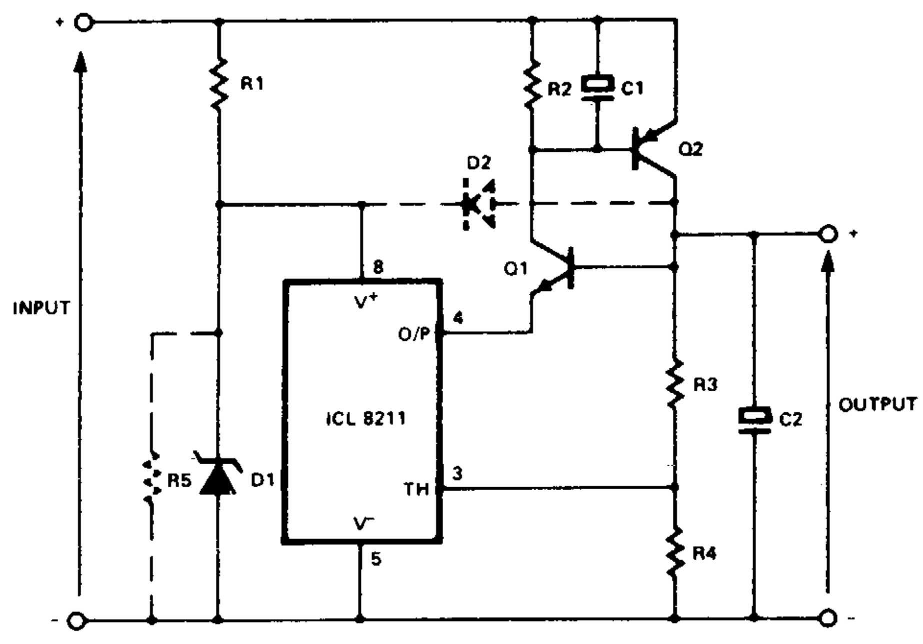


MAXIM

Fig. 73-28

In this circuit, a negative output voltage dc-dc converter generates a −5 V output at pin A. In order to generate −5 V at point A, the primary of the transformer must fly back to a diode drop more negative than −5 V. If the transformer has a tightly coupled 1/1 turns ratio, there will be a 5 V plus a diode drop across the secondary. The 1N5817 rectifies this secondary voltage to generate an isolated 5-V output. The isolated output is not fully regulated since only the −5 V at point A is sensed by the MAX635.

POSITIVE REGULATOR WITH NPN AND PNP BOOST

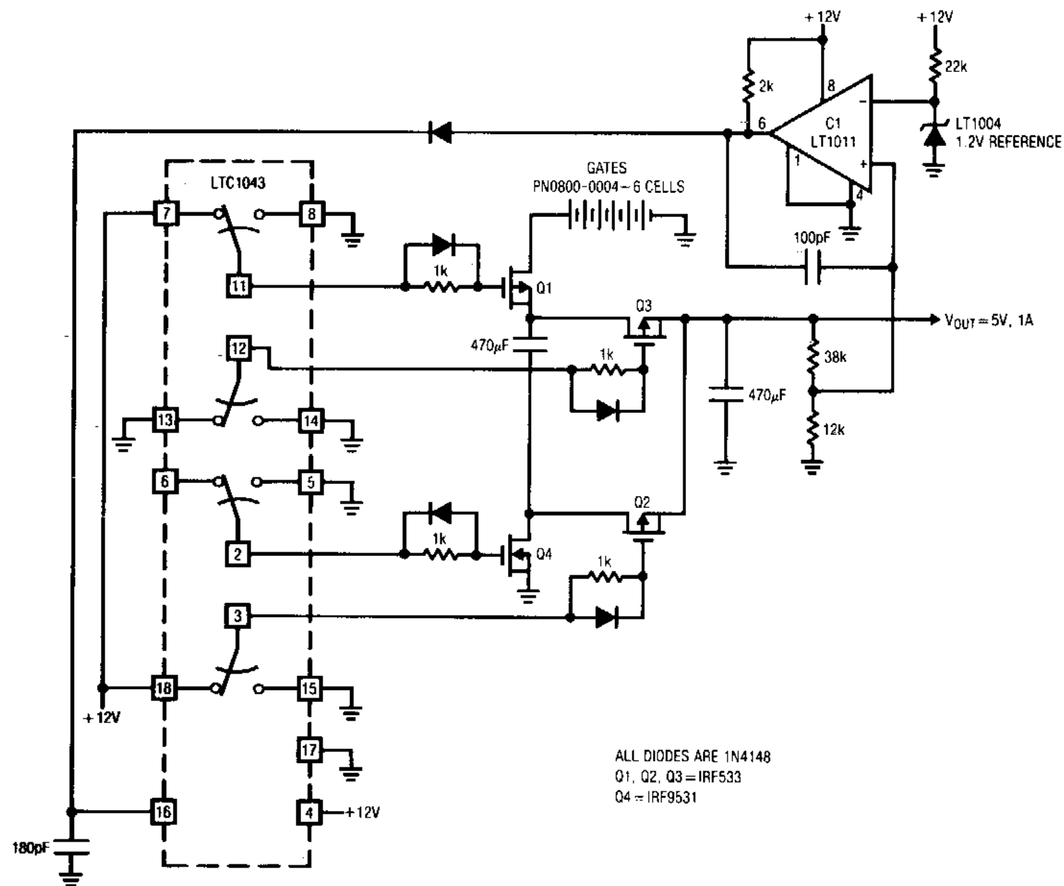


INTERSIL

Fig. 73-29

In the circuit, Q1 and Q2 are connected in the classic SCR or thyristor configuration. Where higher input voltages or minimum component count are required, the circuit for thyristor boost can be used. The thyristor is running in a linear mode with its cathode as the control terminal and its gate as the output terminal. This is known as the remote base configuration.

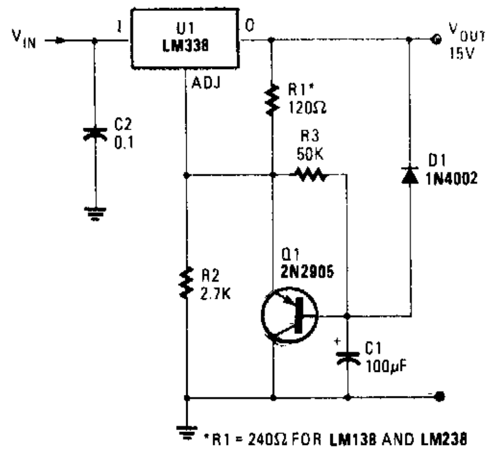
HIGH-CURRENT INDUCTORLESS, SWITCHING REGULATOR



LINEAR TECHNOLOGY CORP. *Fig. 73-30*

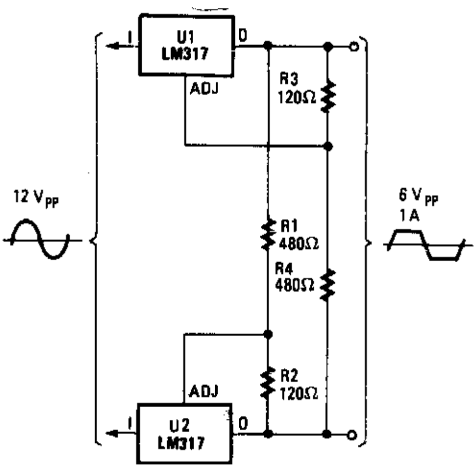
The LTC10432 switched-capacitor building block provides nonoverlapping complementary drive to the Q1 to Q4 power MOSFETs. The MOSFETs are arranged so that C1 and C2 are alternately placed in series and then parallel. During the series phase, the +12 V battery's current flows through both capacitors, charging them, and furnishing load current. During the parallel phase, both capacitors deliver current to the load. Q1 and Q2 receive similar drive from pins 3 and 11. The diode-resistor networks provide additional nonoverlapping drive characteristics, preventing simultaneous drive to the series-parallel phase switches. Normally, the output would be one-half of the supply voltage, but C1 and its associated components close a feedback loop, forcing the output to 5 V. With the circuit in the series phase, the output heads rapidly positive. When the output exceeds 5 V, C1 trips, forcing the LTC1043 oscillator pin, trace D, high; this truncates the LTC1043's triangular-wave oscillator cycle. The circuit is forced into the parallel phase and the output coasts down slowly, until the next LTC1043 clock cycle begins. C1's output diode prevents the triangle down-slope from being affected and the 100-pF capacitor provides sharp transitions. The loop regulates the output to 5 V by feedback controlling the turn-off point of the series phase.

SLOW TURN-ON 15 V REGULATOR



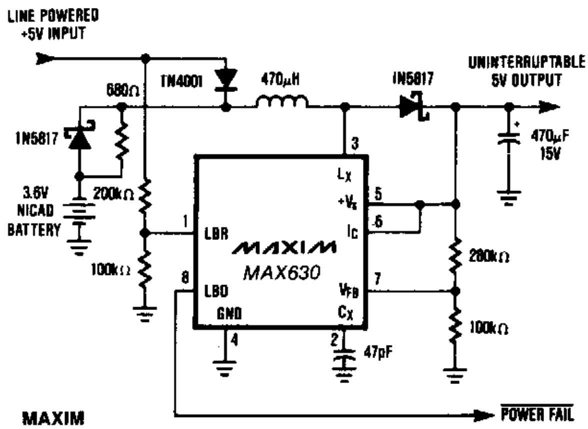
POPULAR ELECTRONICS *Fig. 73-31*

AC VOLTAGE REGULATOR



POPULAR ELECTRONICS *Fig. 73-32*

UNINTERRUPTABLE +5 V SUPPLY



MAXIM *Fig. 73-33*

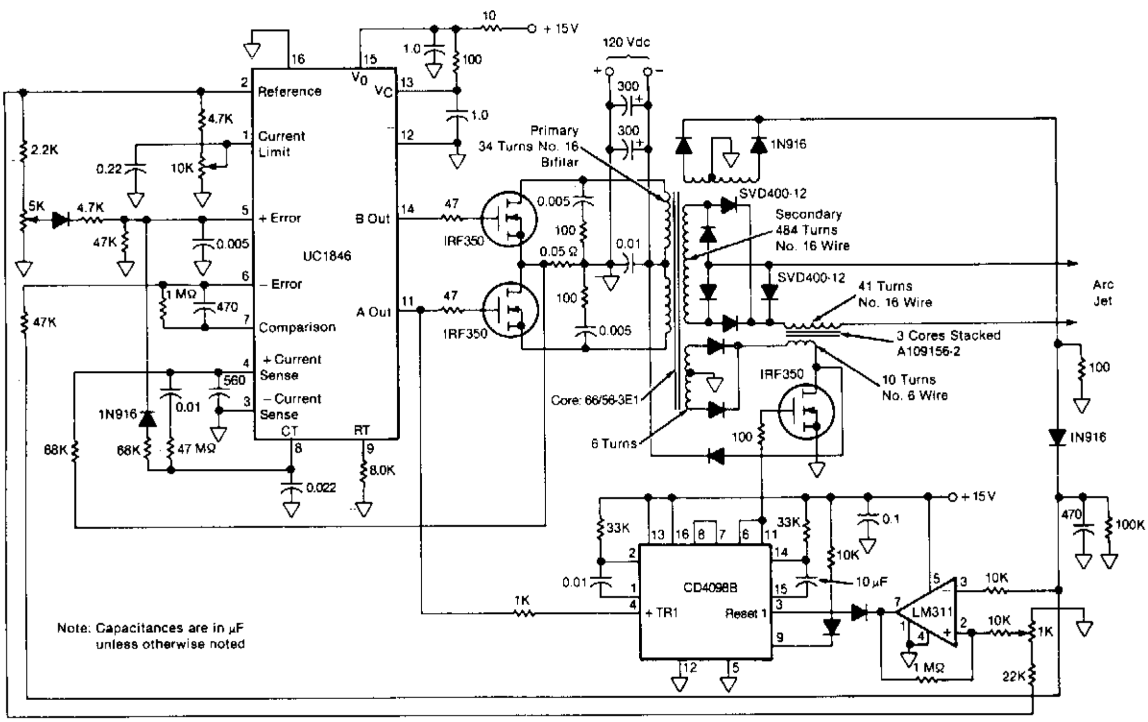
This circuit provides a continuous supply of regulated +5 V, with automatic switch-over between line power and battery backup. When the line-powered input voltage is a +5 V, it provides 4.4 V to the MAX630 and trickle charges the battery. If the line-powered input falls below the battery voltage, the 3.6 V battery supplies power to the MAX630, which boosts the battery voltage up to +5 V, thus maintaining a continuous supply to the uninterruptable +5 V bus. Since the +5 V output is always supplied through the MAX630, there are no power spikes or glitches during power transfer. The MAX630's low-battery detector monitors the line-powered +5 V, and the LBD output can be used to shut down unnecessary sections of the system during power failures. Alternatively, the low-battery detector could monitor the NiCad battery voltage and provide warning of power loss when the battery is nearly discharged. Unlike battery backup systems that use 9-V batteries, this circuit does not need +12 or +15 V to recharge the battery. Consequently, it can be used to provide +5 V backup on modules or circuit cards which only have 5 V available.

74

High-Voltage Power Supplies

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

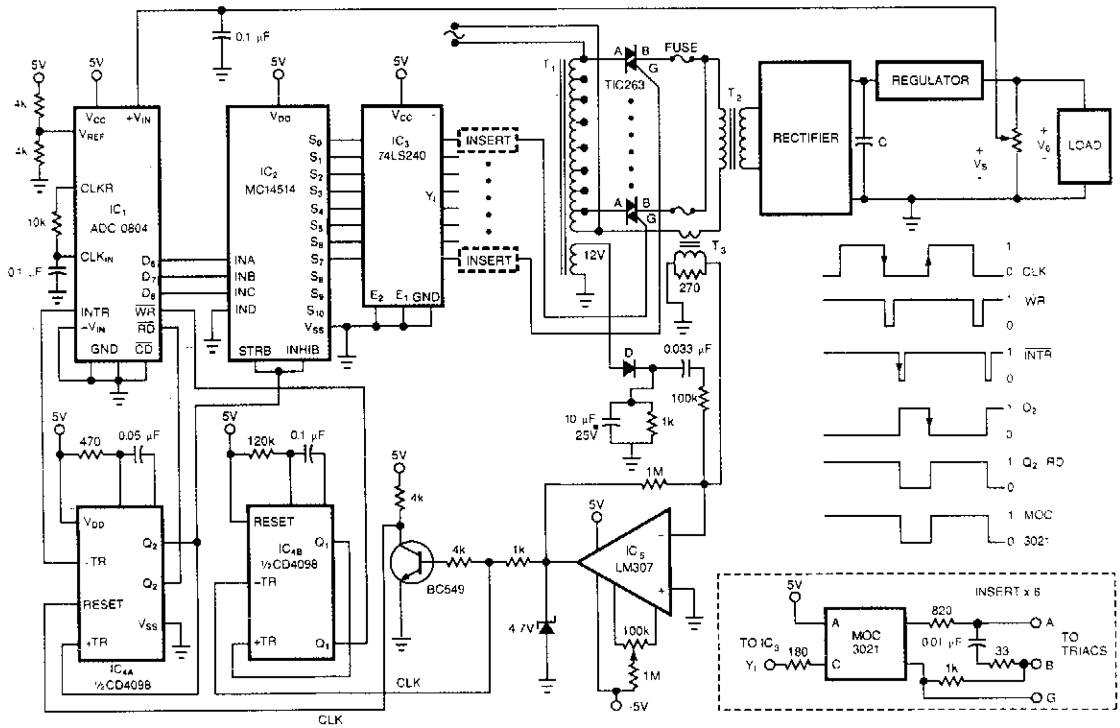
ARC-JET POWER SUPPLY AND STARTING CIRCUIT



NASA TECH BRIEFS *Fig. 74-1*

This circuit for starting arc jets and controlling them in steady operation is capable of high power efficiency and can be constructed in a lightweight form. The design comprises a pulse-width-modulated power converter, which is configured in a closed control loop for fast current control. The series averaging inductor maintains nearly constant current during rapid voltage changes, and thereby allows time for the fast-response regulator to adjust its pulse width to accommodate load-voltage changes. The output averaging inductor doubles as the high-voltage pulse transformer for ignition. The starting circuit operates according to the same principle as that of an automobile ignition coil. When the current is interrupted by a transistor switch, the inductor magnetic field collapses, and a high-voltage pulse is produced. The pulse is initiated every 0.25 second until arc current is detected, then the pulser is automatically turned off.

PREREGULATED HIGH-VOLTAGE SUPPLY



EDN

Fig. 74-2

One of the control circuit's triacs selects the tap on main transformer T1, which provides the proper, preregulated voltage to the secondary regulator. T2 and its associated components comprise the secondary regulator.

The ADC 0804, IC1, digitizes a voltage-feedback signal from the secondary regulator's output. The MC1415 demultiplexer, IC2, decodes the digitizer's output. IC2, in turn, drives T1's optoisolated triacs via the 74LS240 driver chip, IC3, and associated optoisolators.

Transformer T3 samples the circuit's current output. The auxiliary, 12 V winding on T1 ensures no-load starting. The combination of op amp IC5 and the inverting transistor, Q1, square this current signal. The output of Q1 is the CLK signal, which triggers one-half of the one shot, IC4A, to begin the circuit's A/D conversion. The one shots' periods are set to time out within ½ cycle of the ac input.

Upon completion of its A/D conversion, IC1's INTR output triggers the other half of the one shot, IC4B, which enables the converter's data outputs. The rising edge of the CLK signal resets the one shot and latches the new conversion value into IC2. The latch, associated driver, and optoisolator trigger a selected triac according to the latest value of the voltage-feedback signal, V_o.

HIGH-VOLTAGE BUCKING REGULATOR

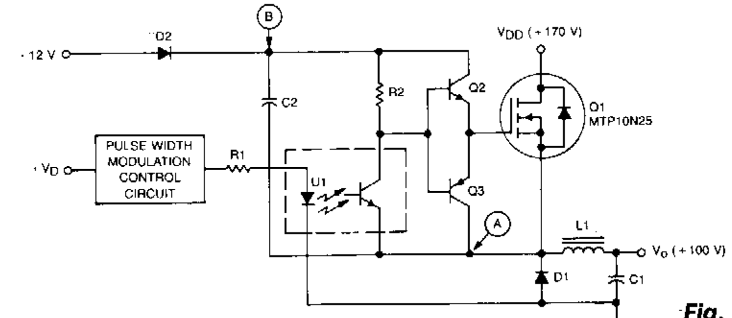


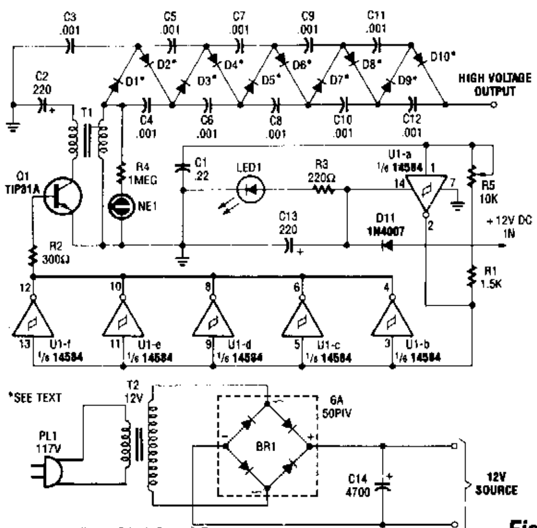
Fig. 74-3

This circuit is basically the classic bucking regulator, except it uses a TMOS N-channel power FET for the chopper and creates its own supply for the gate control.

The unique aspect of this circuit is how it generates a separate supply for the gate circuit, which must be greater than V_{DD}. When power is applied, C2 charges, through D2, to +12 V. At this time, Q1 is off and the voltage at point A is just below zero. When the pulse-modulated signal is applied, the optoisolator transistors, Q2 and Q3, supply a signal to Q1 that turns it on. The voltage at point A then goes to V_{DD}, C2 back-biases D2, and the voltage at point B becomes 12 V above V_{DD}.

After Q1 is turned on, current starts to flow through L1 into C1, increasing until Q1 turns off. The current still wants to flow through L1, so the voltage at point A moves toward negative infinity, but is clamped by D1 to just below zero. Current flows less and less into C1, until Q1 turns on again. Q2 and Q3 drive Q1's gate between the voltages at point A and B, which is always a 12 V swing, so V_{GS} max. is never exceeded. For proper operation, the 12-V supply has to be established before the pulse-width modulator signal is applied.

HIGH-VOLTAGE DC GENERATOR

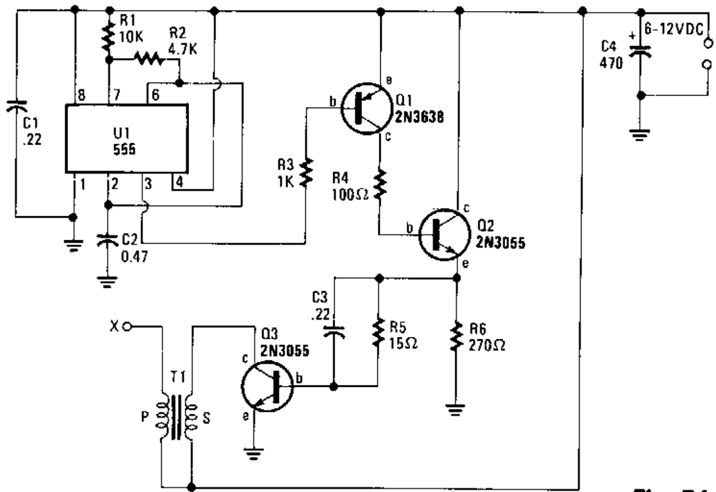


POPULAR ELECTRONICS

Fig. 74-4

This circuit is fed from a 12-Vdc power supply. The input to the circuit is then amplified to provide a 10,000-Vdc output. The output of the up-converter is then fed into a 10 stage, high-voltage multiplier to produce an output of 10,000 Vdc.

BATTERY-POWERED HIGH-VOLTAGE GENERATOR



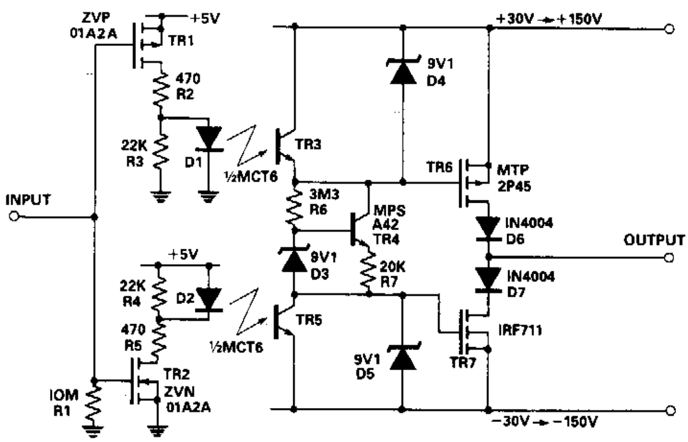
HANDS-ON ELECTRONICS

T1 = Auto Ignition Coil

Fig. 74-5

Output voltage great enough to jump a 1-inch gap can be obtained from a 12-V power source. A 555 timer IC is connected as an astable multivibrator that produces a narrow negative pulse at pin 3. The pulse turns Q1 on for the duration of the time period. The collector of Q1 is direct-coupled to the base of the power transistor Q2, turning it on during the same time period. The emitter of Q2 is direct-coupled through current limiting resistor R5 to the base of the power transistor. Q3 switches on, producing a minimum resistance between the collector and emitter. The high-current pulse going through the primary of high-voltage transformer T1 generates a very high pulse voltage at its secondary output terminal (labeled X). The pulse frequency is determined by the values of R1, R2, and C2. The values given in the parts list were chosen to give the best possible performance when an auto-ignition coil is used for T1.

OPTOISOLATED HIGH-VOLTAGE DRIVER



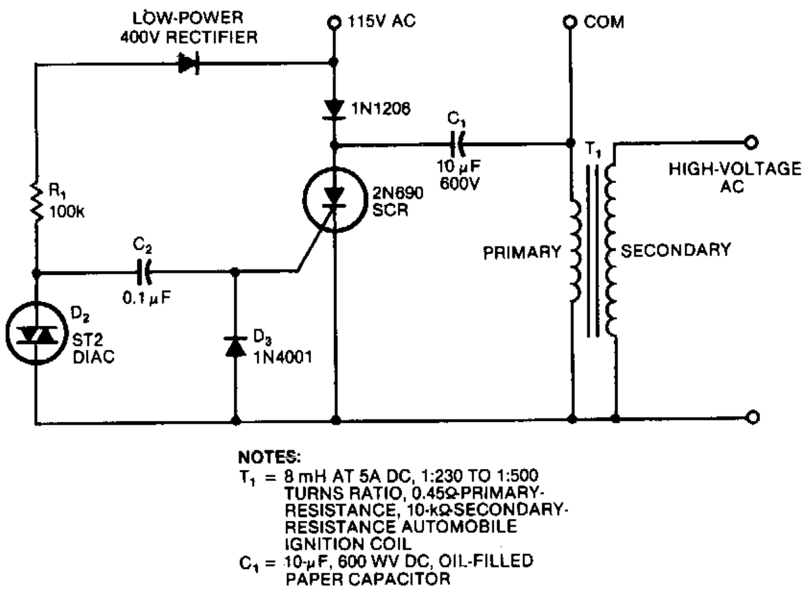
ELECTRONIC ENGINEERING

Fig. 74-6

This circuit takes as an input a signal from a 5-V CMOS logic circuit and outputs a high voltage of the same polarity. The high-voltage supply can be varied from ±30 V to ±150 V without the need to change circuit components. The input voltage is applied to the gates of transistors TR1 and TR2.

TR3 is optically coupled to D1 as is TR5 to D2. R5 limits the current through D2, while R3 and R4 reduce the affects of leakage current. The light transmitted by D1 turns TR3 on and discharges the gate-source capacitance of TR6, which turns TR6 off. At the same time, TR5 is off and a constant current produced by R6, R7, D3, and TR4 charges the gate-sourced capacitance of TR7, thus turning TR7 on. With TR7 on and TR6 off, the output is pulled close to the lower supply rail. When the input is high, TR1 is off and TR2 is on. Therefore, D2 conducts, which turns on TR5. With TR3 off and TR5 on, TR6 turns on and TR7 off. The output is pulled towards the higher supply rail.

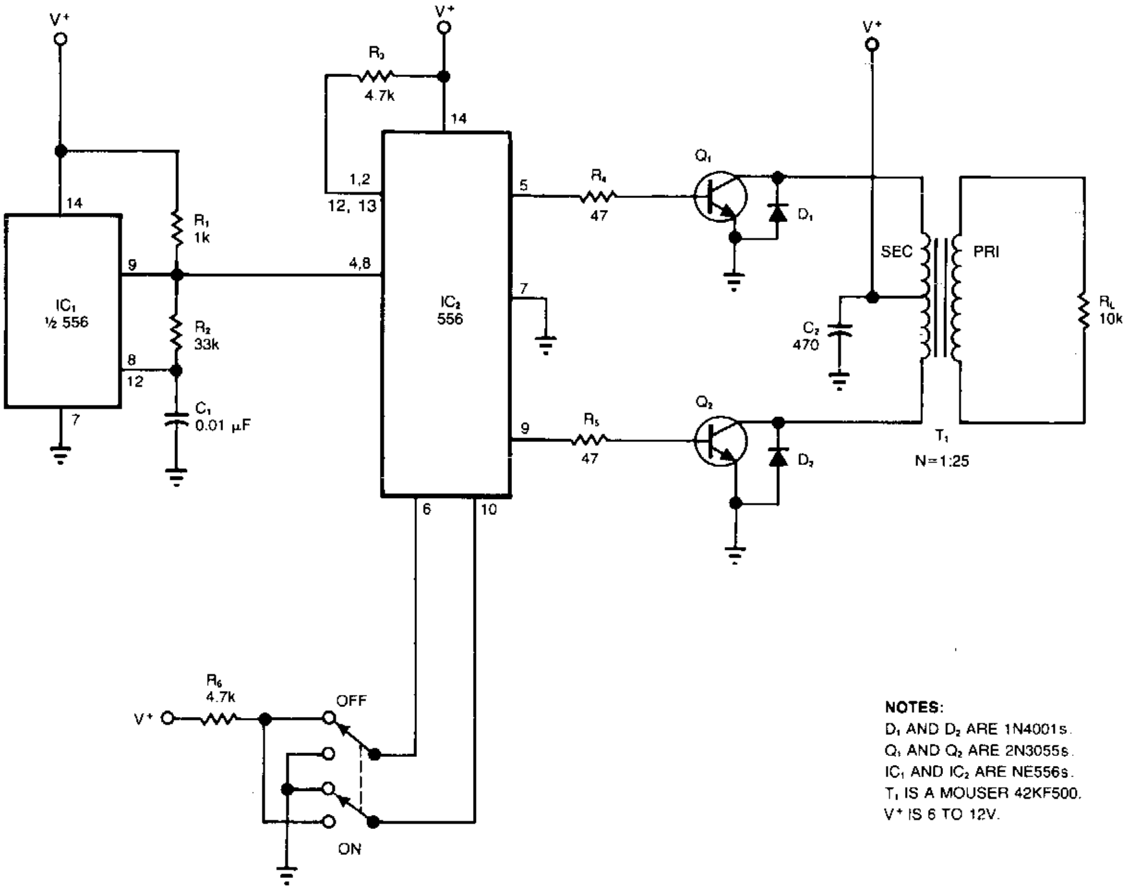
SIMPLE HIGH-VOLTAGE SUPPLY



EDN *Fig. 74-7*

This circuit can generate high-voltage pulses with an inexpensive auto ignition coil. Add a rectifier on the output and the circuit produces high-voltage dc. The circuit's input is 115 Vac. During the input's positive half cycle, energy is stored in capacitor C1, which is charged via diode D1 and the primary winding of transformer T1, the coil. The SCR and its trigger circuitry are inactive during this period. During the input's negative half cycle, energy is stored in capacitor C2 until diac D2 reaches its trigger voltage, whereupon D2 conducts abruptly and C2 releases its energy into the SCR's gate. The SCR then discharges C1 into the transformer's primary and ceases to conduct. This store-and-release cycle repeats on the line's positive and negative half cycles, producing high-voltage pulses at the transformer's secondary.

HIGH-VOLTAGE INVERTER

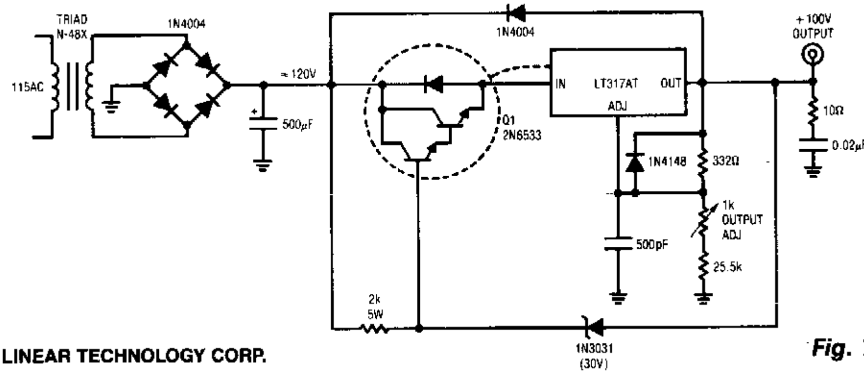


EDN

Fig. 74-8

The circuit converts a dc voltage (V^+) to a high-amplitude square wave in the audio-frequency range. The dual timer, IC2, provides an inexpensive alternative to the traditional transformer for providing complementary base drive to the power transistors, Q1 and Q2. You can convert a 6 to 12 V battery output, for example, to an ac amplitude, which is limited primarily by the power rating of transformer T1. Connect timer IC1 as an oscillator to provide a symmetrical square-wave drive to both inputs of IC2. The timing components, R2 and C1, produce a 2.2-kHz output frequency. By connecting half of IC2 in the inverting mode and the other half in noninverting mode, the timer's outputs alternately drive the two transistors. You can operate the audio-output transformer, T1, as a step-up transformer by connecting it backwards—using the output winding as an input. The transformer delivers an output voltage across R_L of $4 \times N \times V^+$V pk-pk, where N is the transformer turns ratio. For the circuit shown, the output swing is $100 \times V^+$V pk-pk.

HIGH-VOLTAGE REGULATOR



LINEAR TECHNOLOGY CORP.

Fig. 74-9

The regulator delivers 100-V at 100 mA and withstands shorts to ground. Even at 100 V output, the LT317A functions in the normal mode, maintaining 1.2 V between its output and adjustment pin. Under these conditions, the 30-V zener is off and Q1 conducts. When an output short occurs, the zener conducts, forcing Q1's base to 30 V. This causes Q1's emitter to clamp 2 V_{BE}s below V_Z, well within the V_{IN}-V_{OUT} rating of the regulator. Under these conditions, Q1, a high-voltage device, sustains 90 V-V_{CE} at whatever current the transformer specified saturates at 130 mA, while Q1 safely dissipates 12 W. If Q1 and the LT317A are thermally coupled, the regulator will soon go into thermal shutdown and oscillation will commence. This action will continue, protecting the load and the regulator as long as the output remains shorted. The 500-pF capacitor and the 10 Ω/0.02 μF damper aid transient response and the diodes provide safe discharge paths for the capacitors.

CAPACITOR-DISCHARGE HIGH-VOLTAGE GENERATOR

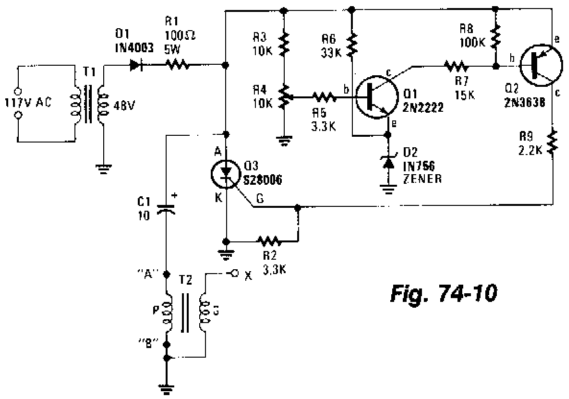
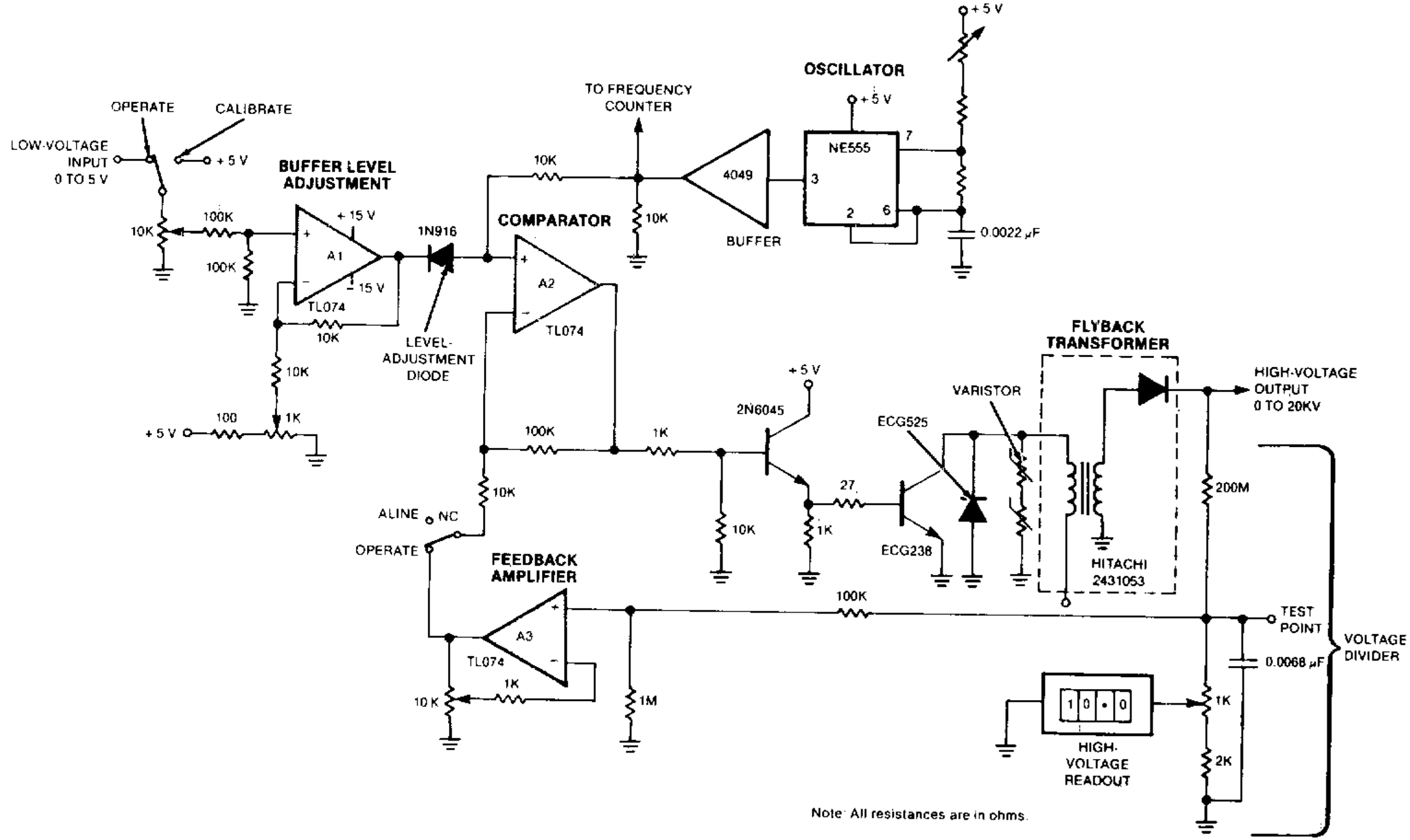


Fig. 74-10

HANDS-ON ELECTRONICS

Stepdown transformer T1 drops the incoming line voltage to approximately 48 Vac which is rectified by diode D1; the resultant dc charges capacitor C1—through current limiting resistor R1—to a voltage level preset by R4. When the voltage on R4's wiper reaches about 8.6 V, Q1 begins to turn on, drawing current through R7 and the base-emitter junction of Q2. Q2 turns on and supplies a positive voltage to the gate of silicon-controlled rectifier Q3. The positive gate voltage causes Q3 to conduct, thereby discharging C1 through the primary winding of step-up transformer T2, which results in a high-voltage arc at output terminal X. The voltage developed at T2's output is determined by the value of C1, the voltage across C1, and the turns ratio of transformer T2. The frequency or pulse rate of the high voltage is determined by the resistance of T1's primary and secondary windings, the value of R1, and the value of C1. The lower the value of each item, the higher the output pulse rate; the peak output voltage will only remain unchanged if C1's value remains unchanged.

REMOTELY ADJUSTABLE SOLID-STATE HIGH-VOLTAGE SUPPLY



NASA

Fig. 74-11

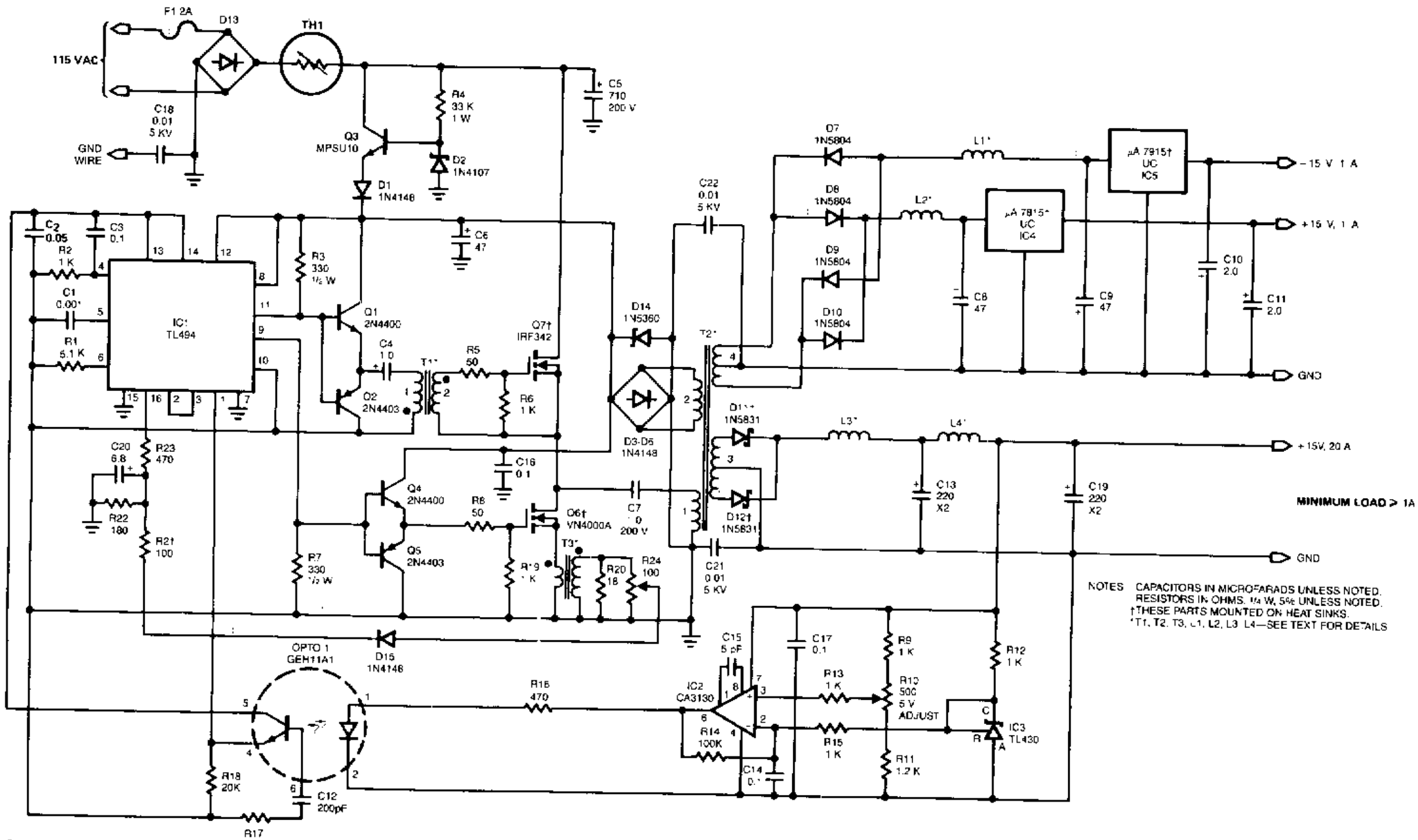
The output voltage changes approximately linearly up to 20 KV as the input voltage is varied from 0 to 5 V. The oscillator is tuned by a 5-Ω potentiometer to peak the output voltage at the frequency of maximum transformer response between 45 and 55 kHz. The feedback voltage is applied through a 100-KΩ resistor, an op amp, and a comparator to a high-voltage amplifier. A diode and varistors on the primary side of the transformer protect the output transistor. The transformer is a flyback-type used in color-television sets. A feedback loop balances between the high-voltage output and the low-voltage input.

75

Variable Power Supplies

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

100-kHz MULTIPLE-OUTPUT SWITCHING POWER SUPPLY

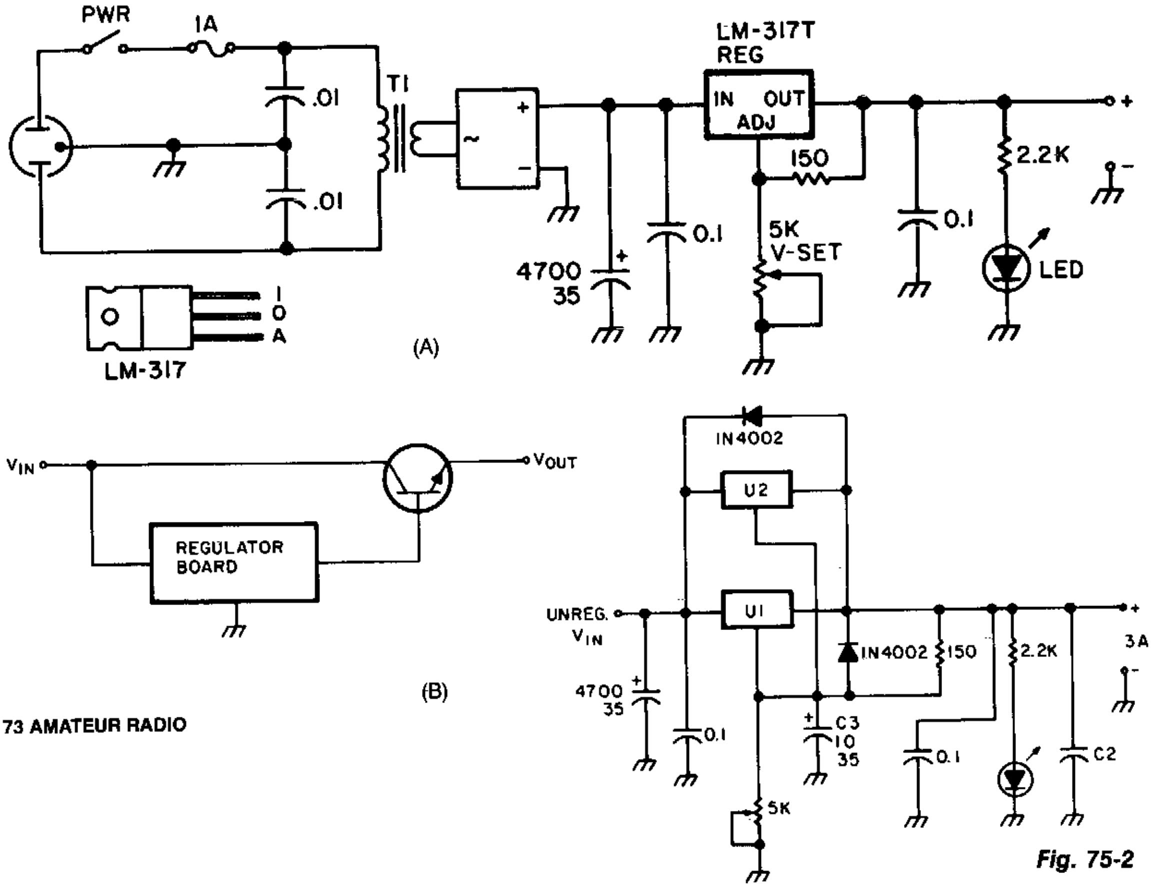


SILICONIX

Fig. 75-1

The power supply uses two VN4000A 400-V MOSPOWER FETs in a half-bridge power switch configuration. Outputs available are +5 V at 20 A and ±15 V (or ±12 V) at 1 A. Since linear three-terminal regulators are used for the low-current outputs, either ±12 V or ±15 V can be made available with a simple change in the transformer secondary windings. A TL494 switching regulator IC provides pulse-width modulation control and drive signals for the power supply. The upper MOSPOWER FET, Q7, in the power switch stage is driven by a simple transformer drive circuit. The lower MOS, Q6, since it is ground referenced, is directly driven from the control IC.

3 – 30 V UNIVERSAL POWER SUPPLY MODULE



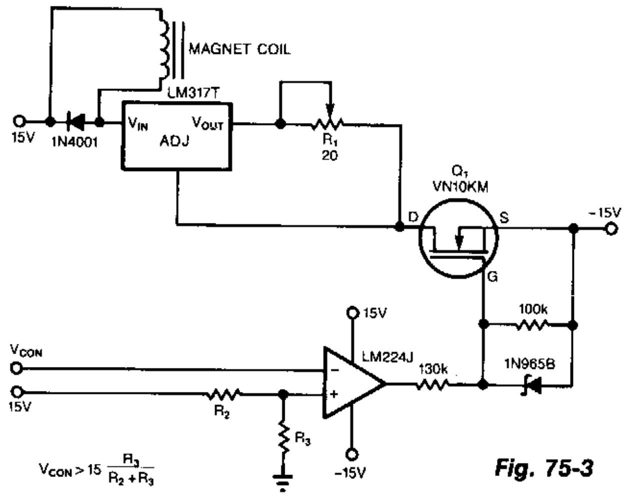
73 AMATEUR RADIO

Fig. 75-2

U1, an LM317 adjustable regulator provides short-circuit protection and automatic current limiting at 1.5 A. The input voltage to the regulator is supplied by DB1, a 4-A 100 PIV full-wave bridge rectifier. Capacitor C1 provides initial filtering. U1 provides additional electronic filtering as part of the regulating function. The output level of the regulator is set by trim-pot R1. Bypass capacitors on the input and output of U1 prevent high-frequency oscillation. The current rating of the transformer must be at least 1.8 times the rated continuous-duty output of the supply. This means that a 1.5-A supply should use a 2.7-A transformer. For light or intermittent loads, a smaller 2.0-A transformer should suffice.

Wiring a second LM317, U2, in parallel with U1 is a quick and clean way to increase the current-limiting threshold to 3 A without sacrificing short-circuit protection. When more than 3 A is required, the regulator module can be used to drive the base of one or more pass-transistors (see Fig. 75-2B).

REGULATOR/CURRENT SOURCE

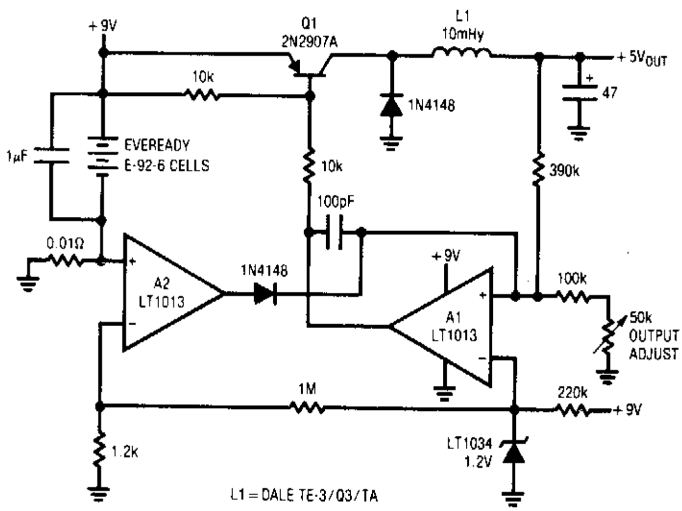


EDN

Fig. 75-3

The circuit powers the load via the regulator's input instead of its output. Because the regulator's output sees constant dummy load R1, it tries to consume a constant amount of current, no matter what the voltage across the actual load really is. Hence, the regulator's input serves as a constant-current source for the actual load. Power the circuit with any one of the commonly available ±15 or ±12 V supplies. The voltage dropped across the regulator and dummy load decreased the total compliance voltage of the circuit. You set the load's current with R1. The current equals 1.25 A/Ω × *R*1.

LOW-POWER SWITCHING REGULATOR

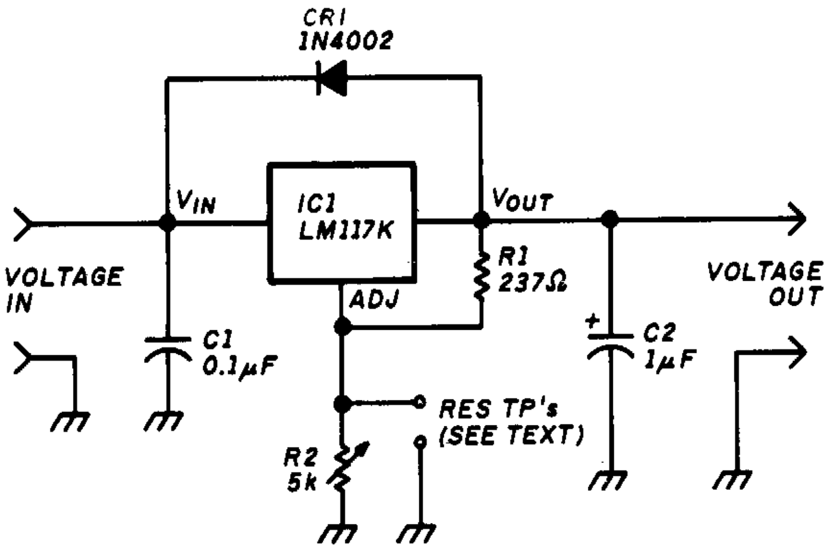


LINEAR TECHNOLOGY CORP.

Fig. 75-4

A simple battery-powered switching regulator provides 5 V out from a 9-V source with 80% efficiency and 50 mA output capability. When Q1 is on, its collector voltage rises, forcing current through the inductor. The output voltage rises, causing A1's output to rise. Q1 cuts off and the output decays through the load. The 100-pF capacitor ensures clean switching. The cycle repeats when the output drops low enough for A1 to turn on Q1. The 1-μF capacitor ensures low battery impedance at high frequencies, preventing *sag* during switching.

VARIABLE VOLTAGE REGULATOR

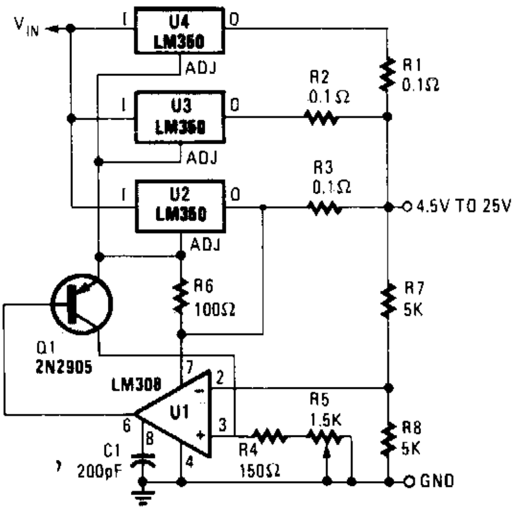


HAM RADIO

Fig. 75-5

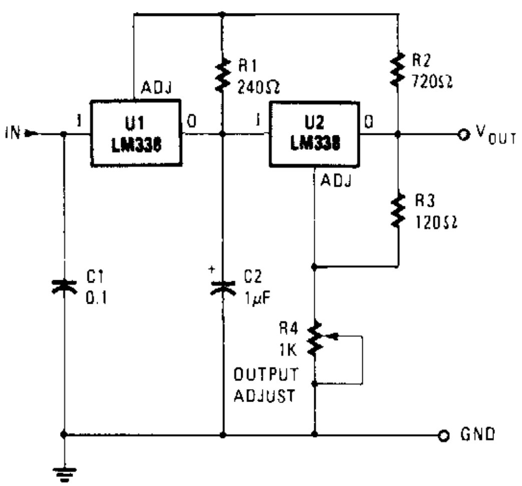
The variable voltage regulator lets you adjust the output voltage of a fixed dc power supply between 1.2 and 37 Vdc, and will supply the output current in excess of 1.5 A. The circuit incorporates an LM117K three-terminal adjustable output positive voltage regulator in a TO-3 can. Thermal overload protection and short-circuit current-limiting constant with temperature are included in the package. Capacitor C1 reduces sensitivity to input line impedance, and C2 reduces excessive ringing. Diode CR1 prevents C2 from discharging through the IC during an output short.

TRACKING PREREGULATOR



POPULAR ELECTRONICS *Fig. 75-6*

ADJUSTABLE 10-A REGULATOR



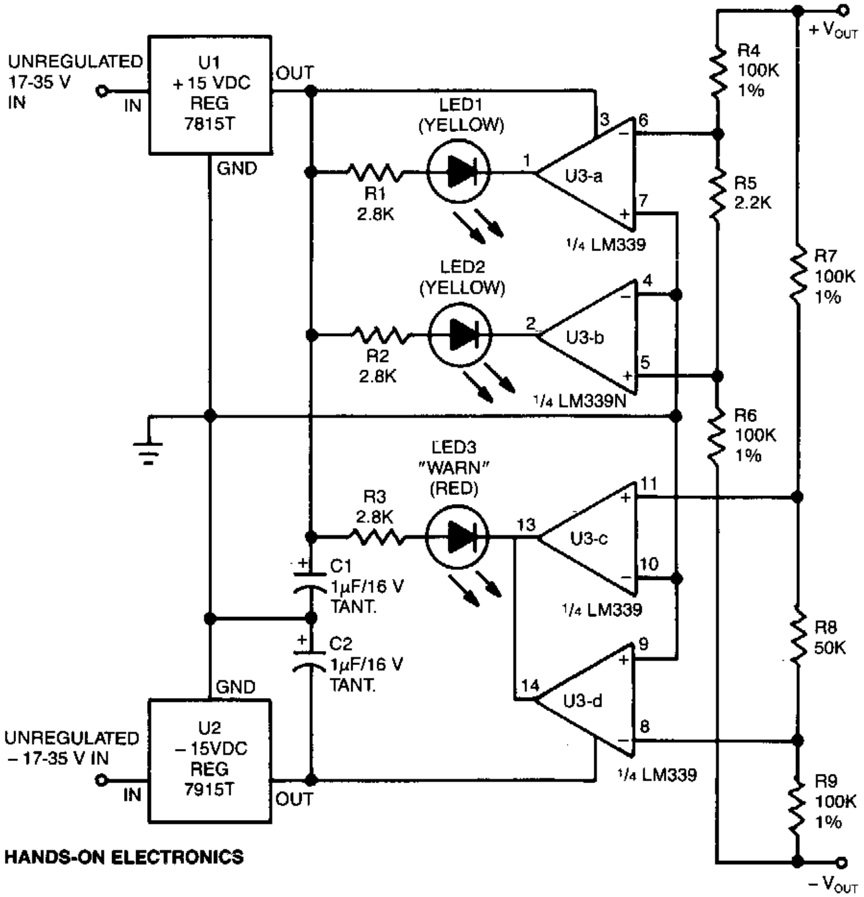
POPULAR ELECTRONICS *Fig. 75-7*

76

Power Supply Monitors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

POWER-SUPPLY BALANCE INDICATOR

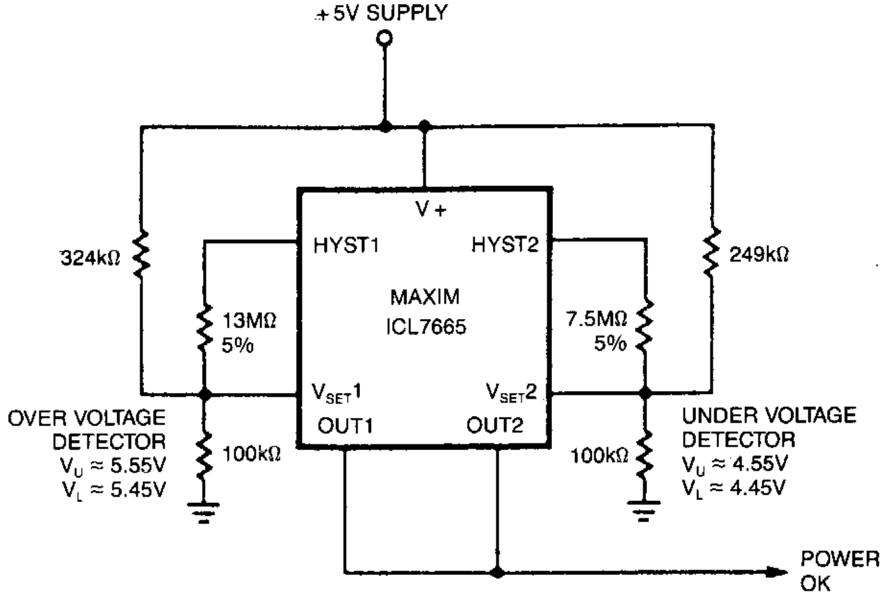


HANDS-ON ELECTRONICS

Fig. 76-1

This circuit uses two comparator pairs from an LM339N quad comparator; one pair drives the yellow positive (+) and negative (−) indicators, the other jointly drives the red warn LED3. The circuit draws its power from the unregulated portion of the power supply. The four comparators get their switching inputs from two parallel resistor-divider strings. Both strings have their ends tied between the power supply's positive and negative output terminals. The first string, consisting of R4, R5, and R6, divides the input voltage in half, with output taps at 0.5%. The other string, made up of R7, R8, and R9, also divides the input voltage in half, with taps at +10%. The 0.5% R4/R5/R6 string drives the two comparators controlling the positive and negative indicators (LED1 and LED2). Their inputs are crossed so that LED2 does not fire until the positive supply is at least 0.5% higher than the negative; the positive indicator does not go off until the negative supply is at least 0.5% higher than the positive—in relative levels. That overlap permits both LEDs to be on when the two supplies are in 1% or better balance. The +10T R7/R8/R9 string drives the other two comparators, which control the warn indicator. If either side of the supply is 10% or more higher than the other, one of the two comparators will switch its output low and light the red LED3—the LM339N has opened-collector outputs, allowing such wired OR connections. The inputs are not crossed, as with the other comparator pair, so there is a band in the middle where neither comparators output is low and the LED remains off.

SINGLE-SUPPLY FAULT MONITOR



MAXIM

Fig. 76-2

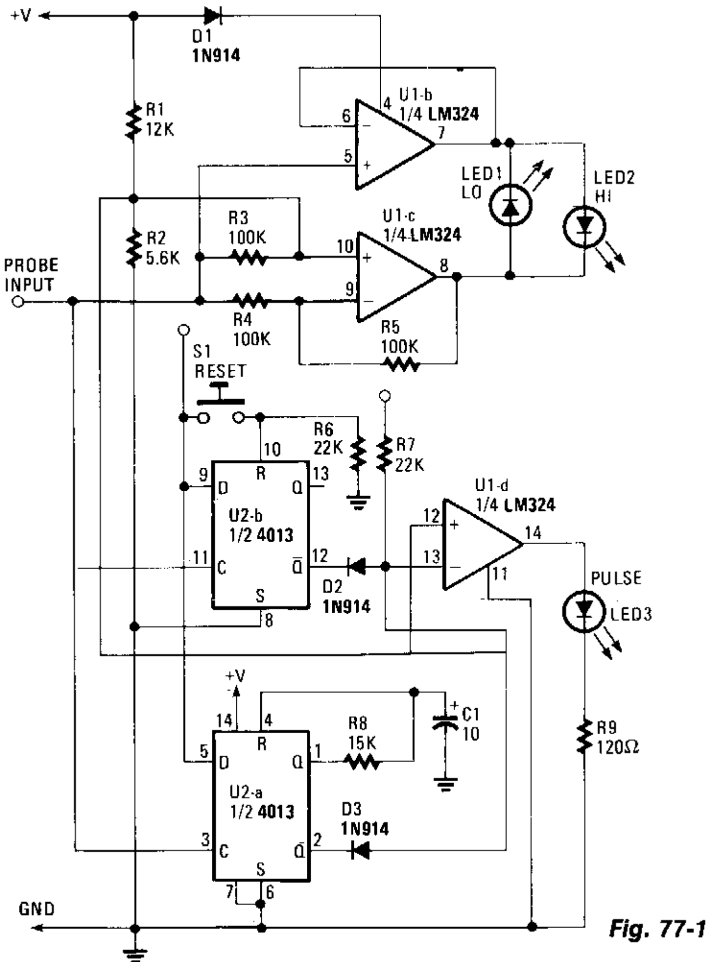
This circuit shows a typical over/under-voltage fault monitor for a single supply. The upper trip points, controlling OUT 1, are centered on 5.5 V with 100 mV of hysteresis (V_U = 5.55 V, V_L = 5.45 V); and the lower trip points, controlling OUT 2, are centered on 4.5 V, also with 100 mV of hysteresis. OUT 1 and OUT 2 are connected together in a wired OR configuration to generate a *power OK* signal.

77

Probes

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DIGITAL LOGIC PROBE



HANDS-ON ELECTRONICS

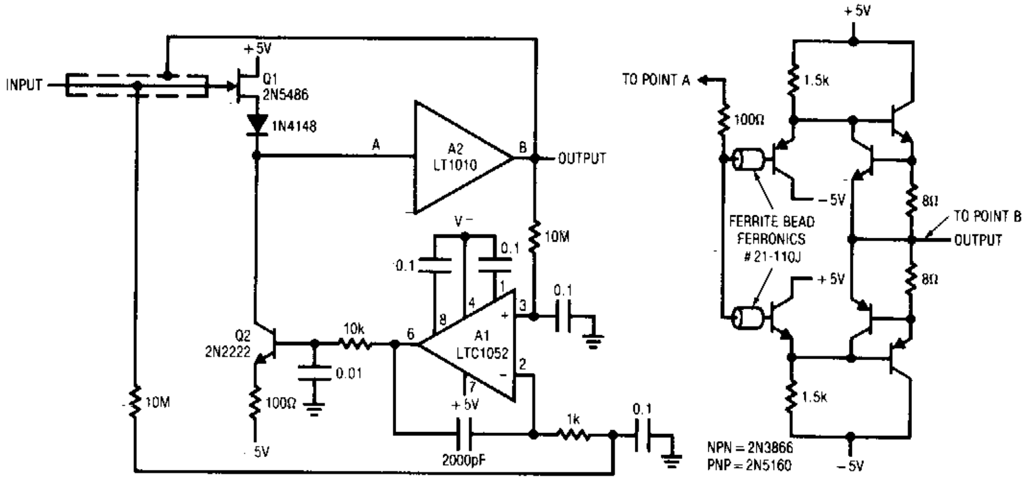
Fig. 77-1

The probe relies on the power supply of the CUT (circuit-under-test). The input to the probe, at probe tip, is fed along two paths. One path flows to the clock inputs of U2a and U2b. The other path feeds both the inverting input of U1c, which is set up as an inverting-mode integrator, and the noninverting input of U1b, which is configured as a noninverting unity-gain amplifier, in a logic-low state.

That low, below the reference set at pin 10, causes U1b's output at pin 7 to become high. With Ub1 outputting low and U1c outputting high, LED1 is forward-biased, and lights. LED2, reverse-biased, remains dark. Suppose that the logic level on the same pin becomes high. That high is applied to pin 5 of U1b, causing its output to be high. LED2 is now forward-biased and lights, while LED1 is reverse-biased and becomes dark.

Assume that a clock frequency is sensed at the probe input; LED1 and LED2 alternately light, and depending on the frequency of the signal, can appear constantly lit. That frequency, which is also applied to the clock input of both flip-flops, causes the *Q* outputs of U2a and U2b to simultaneously alternate between high and low. Each time that the *Q* outputs of the two flip-flops decrease, the output of U1d increases, lighting LED3, indicating that a pulse stream has been detected.

LOW INPUT CAPACITANCE BUFFER

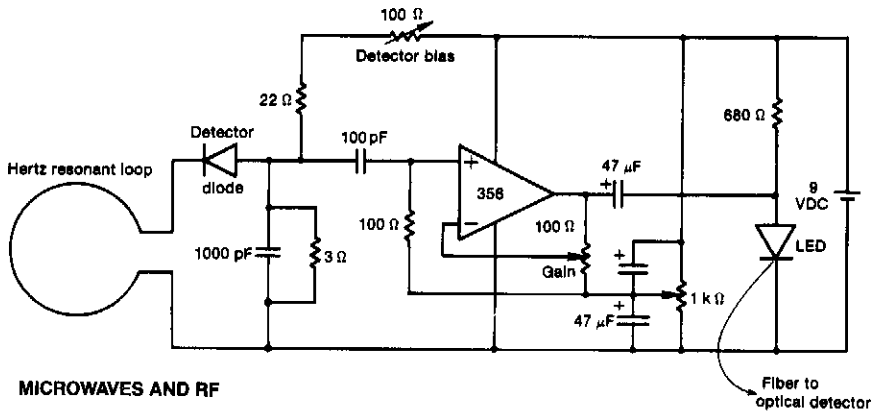


LINEAR TECHNOLOGY CORP.

Fig. 77-2

Q1 and Q2 constitute a simple, high-speed FET input buffer. Q1 functions as a source follower, with the Q2 current-source load setting the drain-source channel current. The LT1010 buffer provides output drive capability for cables or whatever load is required. The LTC1052 stabilizes the circuit by comparing the filtered circuit output to a similarly filtered version of the input signal. The amplified difference between these signals is used to set Q2's bias, and hence Q1's channel current. This forces Q1's V_{GS} to whatever voltage is required to match the circuit's input and output potentials. The diode in Q1's source line ensures that the gate never forward biases and the 2000-pF capacitor at A1 provides stable loop compensation. The rc network in A1's output prevents it from seeing high-speed edges coupled through Q2's collector-base junction. A2's output is also fed back to the shield around Q1's gate lead, bootstrapping the circuit's effective input capacitance to less than 1 pF.

RF PROBE



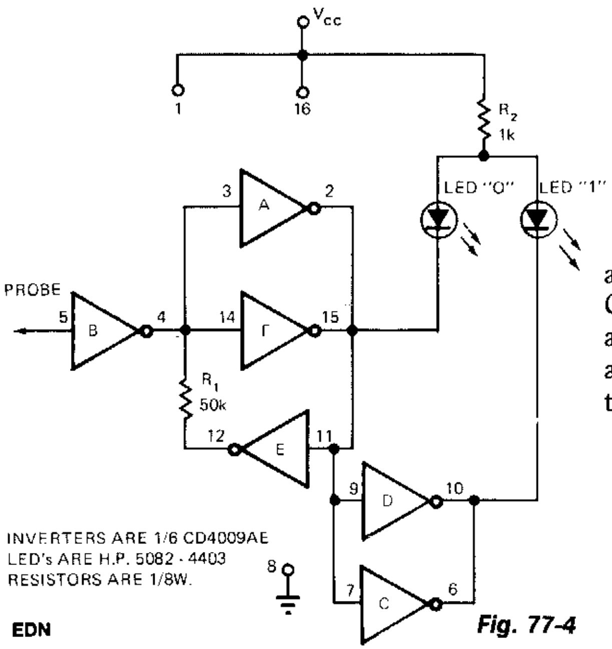
MICROWAVES AND RF

Fig. 77-3

RF PROBE (*Cont.*)

This rf probe is coupled with a fiber-optic cable to the test equipment. It utilizes inexpensive components to improve probe performance at UHF frequencies. The receiving antenna in this probe feeds an envelope-detector diode. After amplification by the LF356 op amp, the low-frequency output modulates the LED, which in turn feeds the optical fiber. The design facilitates the use of a single battery for the op amp, with voltage splitting by means of the 1-KΩ potentiometer, and miniature 47-μF tantalum capacitors to provide decoupling. The gain control is easily adjusted to give the best dynamic range for a specific LED.

CMOS UNIVERSAL LOGIC PROBE

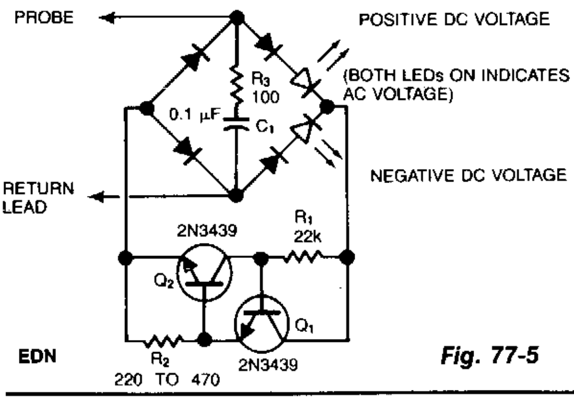


EDN

Fig. 77-4

Only the CD4009AE hex buffer, two resistors, and two LEDs are required for a logic probe. CMOS logic probe features $10^{12}\Omega$ input impedance and covers 3 to 15 V range. While LEDs are visible at all voltages, a 1-KΩ pot in place of R2 will allow the user to increase brightness at lower voltages.

4 – 220 V TEST PROBE

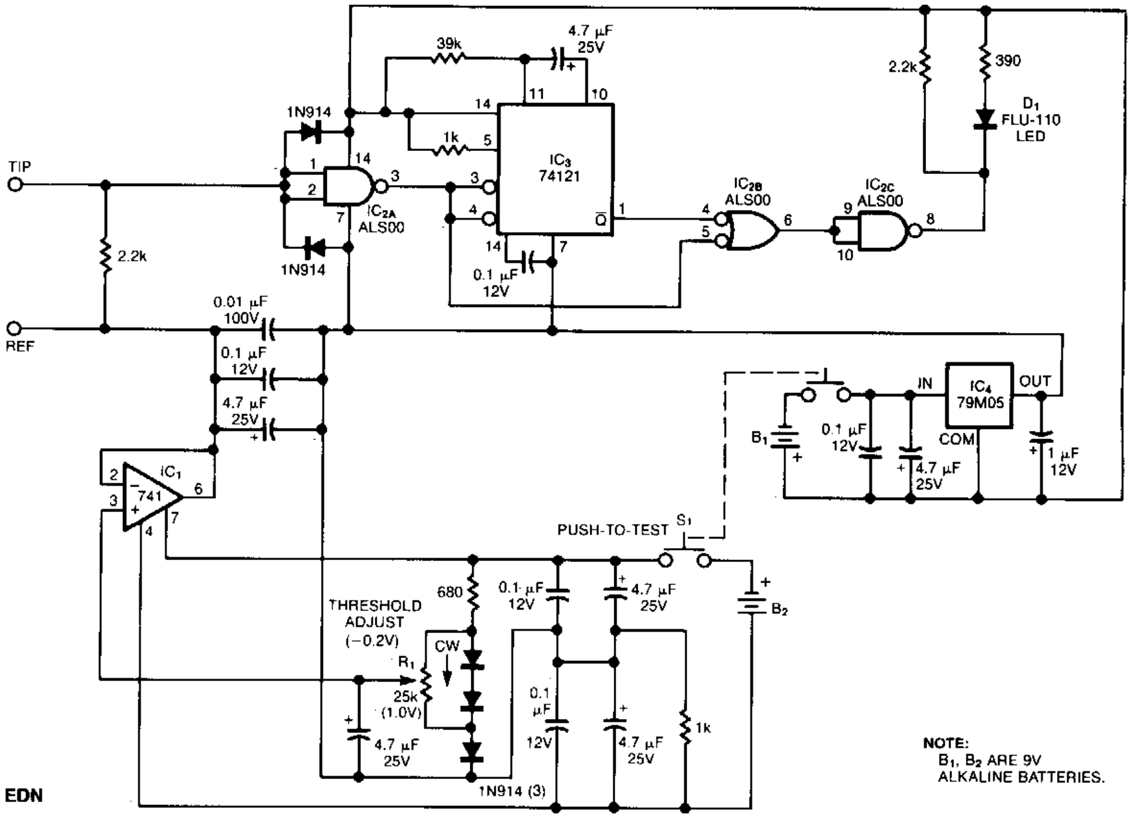


EDN

Fig. 77-5

Using inexpensive components, you can fit a simple probe circuit into a pencil-sized enclosure. When both LEDs are on, the probe indicates the presence of an ac voltage; either LED alone indicates the presence and polarity of a dc voltage. The diode-bridge arrangement allows one-way current source R1, R2, Q1, and Q2 to light either LED (or both) when the probe is activated by a test voltage. Diodes, provide the necessary peak-inverse voltage rating; R3 and C1 provide a spike-suppression network to protect the current-source transistors.

BATTERY-POWERED GROUND-NOISE PROBE

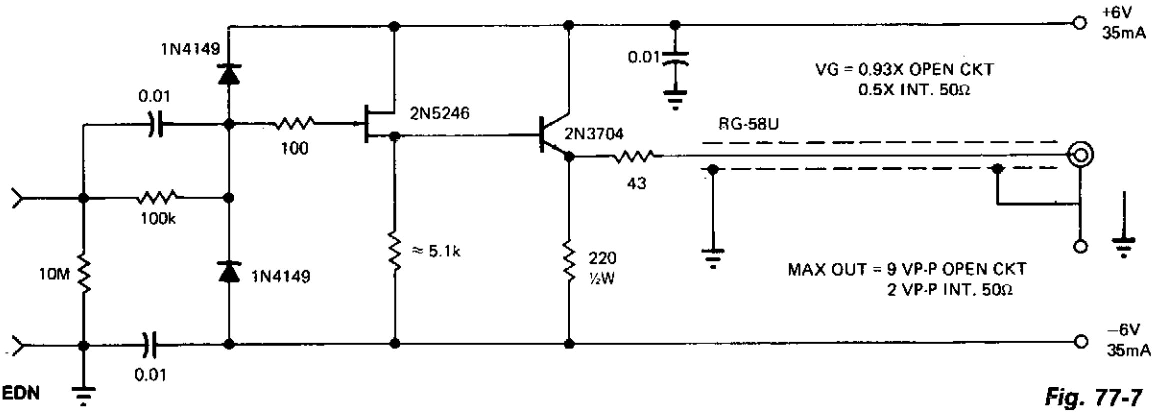


EDN

Fig. 77-6

Oscilloscope measurements of ground noise can be unreliable because noise can enter your circuit via the scope's three-pronged power plug. You can avoid this problem by using the ground-noise tester shown. Powered by two 9-V batteries, the circuit dissipates power only while push-to-test switch S1 is depressed. Noise pulses that reach IC2A's switching threshold of about 1.5 to 1.8 V create a logic transition that triggers the monostable multivibrator IC3, which stretches the pulse to produce a visible blink from LED D1. You set the noise reference level by adjusting threshold-adjust potentiometer R1, which lets the circuit respond to minimum pulse amplitudes ranging from about 0 to 1 V. For convenience, you can use a one-turn potentiometer for R1 and calibrate the dial by applying an adjustable dc voltage, monitored by an accurate voltmeter.

FET PROBE

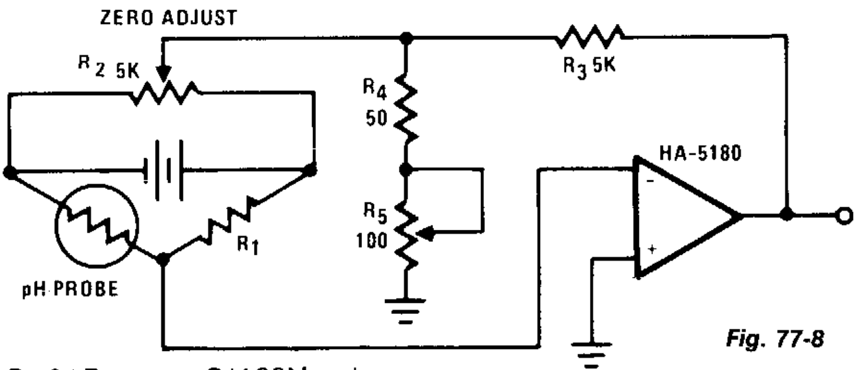


EDN

Fig. 77-7

This FET probe has an input impedance of 10 MΩ shunted by 8 pF. Eliminating the protective diodes reduces this impedance to about 4 pF. The frequency response of the probe extends from dc to 20 MHz (−1 dB), although higher frequency operation is possible through optimized construction and use of a UHF-type transistor. Zero dc offset at the output is achieved by selecting a combination of a 2N5246 and source resistor that yields a gate-source bias equal to the V_{BE} of the 2N3704 at approximately 0 V. At medium frequencies, the probe can be used unterminated for near-unity gain; for optimum impedance converter probe high-frequency response, the cable must be terminated into 50 Ω. The voltage gain, when properly terminated, is precisely 0.5 *X*.

pH PROBE AND DETECTOR



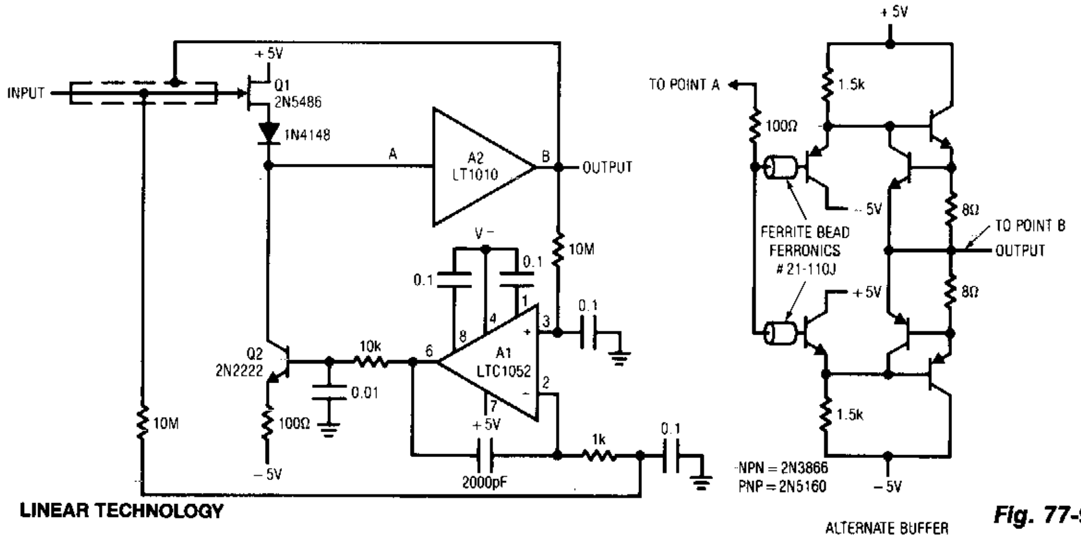
HARRIS

Fig. 77-8

$R_1 \approx R_{PROBE} \approx 100$Megohm

The greatest sensitivity is achieved if R1 is approximately equal to the probe resistance. The circuit can be *zeroed* with R2, while the full-scale voltage is controlled by R5. The correlation between pH and output voltage might not be linear, which would necessitate a shaping circuit. A calibration scheme, using solutions of known pH, might prove adequate and more reliable over a period of time because of probe variance.

STABILIZED LOW-INPUT CAPACITANCE BUFFER

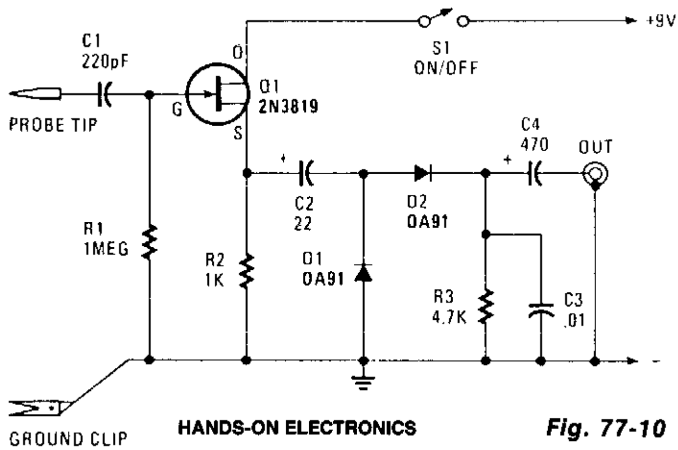


LINEAR TECHNOLOGY

Fig. 77-9

Q1 and Q2 constitute a simple, high-speed FET input buffer. Q1 functions as a source follower, with the Q2 current source load setting the drain-source channel current. The LT1010 buffer provides output drive capability for cables or whatever load is required. Normally, this open-loop configuration would be quite drifty because there is no dc feedback. The LTC1052 contributes this function to stabilize the circuit. It does this by comparing the filtered circuit output to a similarly filtered version of the input signal. The amplified difference between these signals is used to set Q2's bias, and hence Q1's channel current. Q1's source line ensures that the gate never forward biases, and the 2000 pF capacitor at A1 provides stable loop compensation. The rc network in A1's output prevents it from seeing high-speed edges coupled through Q2's collector-base junction. A2's output is also fed back to the shield around Q1's gate lead, bootstrapping the circuit's effective input capacitance down to less than 1 pF.

RF PROBE



HANDS-ON ELECTRONICS

Fig. 77-10

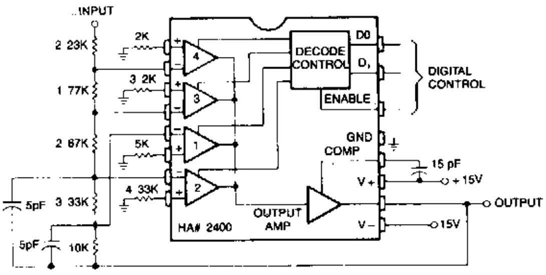
Transistor Q1-configured as a source-follower buffer stage, offering a bit under unity voltage gain—gives the unit a high-impedance input of about 1 MΩ shunted by about 10 pF, which keeps only minimal loading on the equipment being tested. C1 serves as input dc blocking capacitor. The Q1 output is coupled by C2 to a simple AM detector circuit made up of D1, D2, R3 and C3. Capacitor C4 provides output dc blocking. Total current consumption should be somewhere around 1 mA. The circuit responds to frequencies from 100 kHz to well over 50 MHz.

78

Programmable Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

INVERTING PROGRAMMABLE-GAIN AMPLIFIER

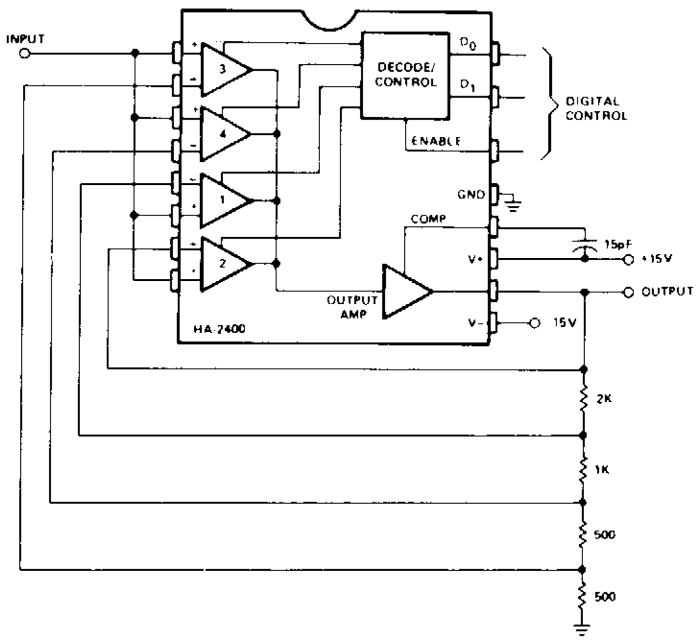


HARRIS

Fig. 78-1

This circuit can be programmed for a gain of 0, −1, −2, −4, or −8. This could also be accomplished with one input resistor and one feedback resistor per channel in the conventional manner, but this would require eight resistors, rather than five.

NONINVERTING PROGRAMMABLE GAIN AMPLIFIER

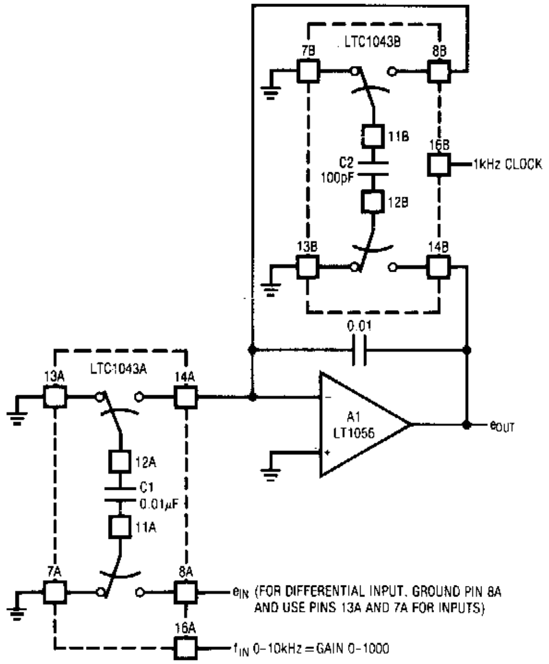


HARRIS

Fig. 78-2

This is a noninverting amplifier configuration with feedback resistors chosen to produce a gain of 0, 1, 2, 4, or 8, depending on the digital control inputs. Comparators at the output could be used for automatic gain selection for auto-ranging meters, etc.

WIDE RANGE DIGITALLY CONTROLLED VARIABLE-GAIN AMPLIFIER

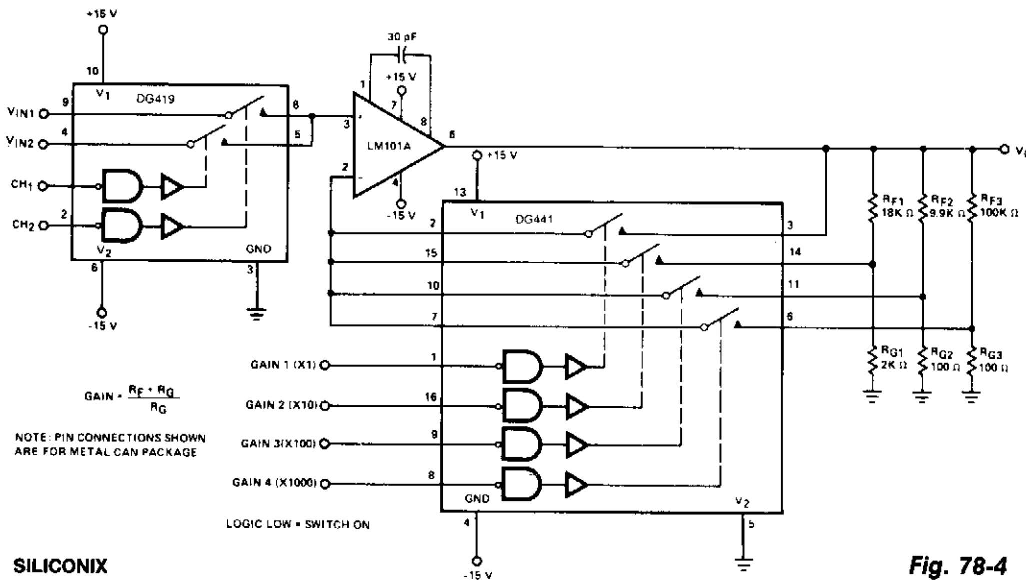


The circuit uses the LTC1043 in a variable gain amplifier which features continuously variable gain, gain stability of 20 ppm/°C, and single-ended or differential inputs. The circuit uses two separate LTC1043s. LTC1043B is continuously clocked by a 1-kHz source, which could also be processor supplied. Both LTC1043s function as the sampled data equivalent of a resistor within the bandwidth set by A1's 0.01-μF value and the switched-capacitor equivalent feedback resistor. The time-averaged current delivered to the summing point by LTC-1043A is a function of the 0.01-μF capacitor's input-derived voltage and the commutation frequency at pin 16. Low-commutation frequencies result in small time-averaged current values, and require a large input resistor. Higher frequencies require an equivalent small input resistor.

LINEAR TECHNOLOGY CORP.

Fig. 78-3

DIGITALLY PROGRAMMABLE PRECISION AMPLIFIER

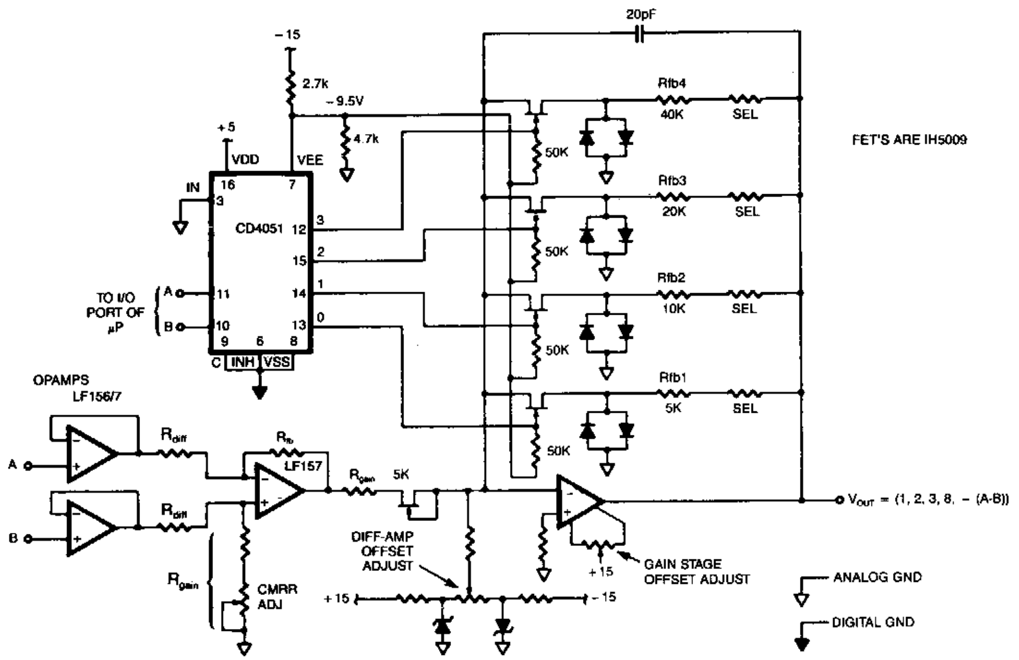


SILICONIX

Fig. 78-4

The DG419 *looks* into the high input impedance of the op amp, so the effects of $R_{DS(on)}$ are negligible. The DG441 is also connected in series with R_{IN} and is not included in the feedback dividers, thus contributing negligible error to the overall gain. Because the DG419 and DG441 can handle ±15 V, the unity gain follower connection, X1, is capable of the full op-amp output range of ±12 V.

PROGRAMMABLE-GAIN DIFFERENTIAL-INPUT AMPLIFIER



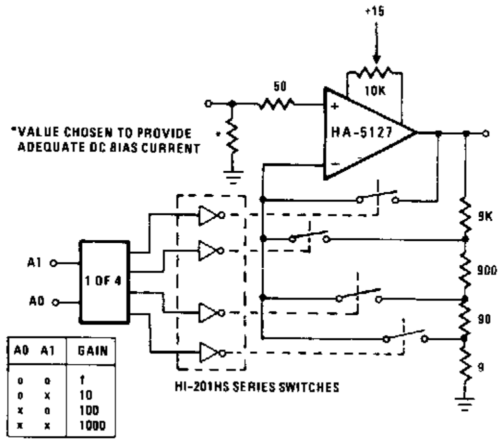
b	a	Gain
0	0	X1
0	1	X2
1	0	X4
1	1	X8

INTERSIL

Fig. 78-5

This programmable gain circuit employs a CD4051 CMOS Analog Multiplexer as a two to four line decoder, with appropriate FET drive for switching between feedback resistors to program the gain to any one of four values.

PROGRAMMABLE AMPLIFIER



HARRIS *Fig. 78-6*

Often a circuit will be called upon to perform several functions. In these situations, the variable gain configuration of this circuit could be quite useful. This programmable gain stage depends on CMOS analog switches to alter the amount of feedback, and thereby, the gain of the stage. Placement of the switching elements inside the relatively low-current area of the feedback loop, minimizes the effects of bias currents and switch resistance on the calculated gain of the stage. Voltage spikes can occur during the switching process, resulting in temporarily reduced gain because of the make-before-break operation of the switches. This gain loss can be minimized by providing a separate voltage divider network for each level of gain.

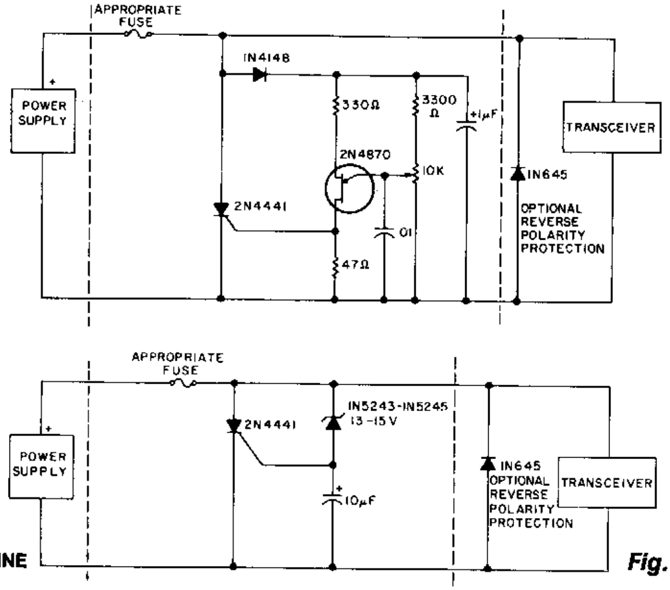
79

Protection Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Electric Crowbars
Ac Power-Line Connections Monitor
Line-Voltage Monitor
Power-Failure Alarm
Ac Circuit Breaker
Fast Overvoltage Protector

ELECTRIC CROWBARS

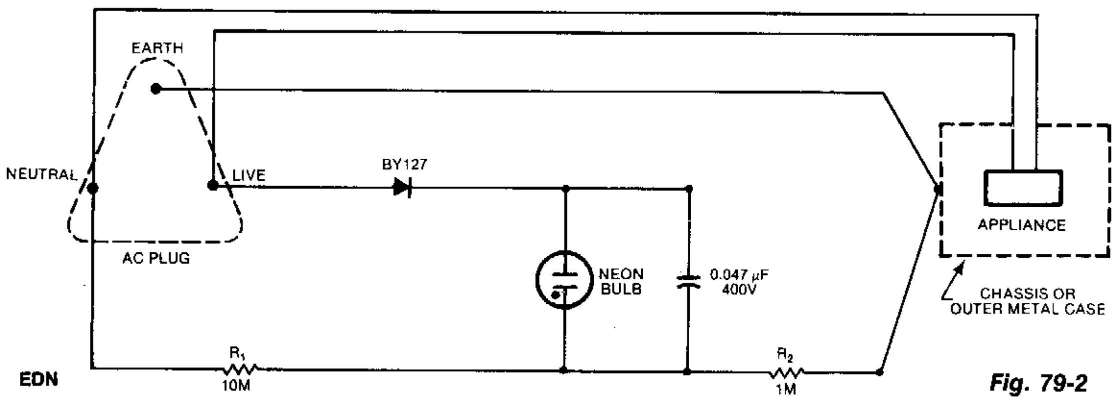


73 MAGAZINE

Fig. 79-1

To avoid grief when using 12-V power supplies with mobile transceivers, especially when there is a short-circuit failure of the series pass transistor, crowbar circuits provide protection by clamping the power line and blowing the fuse within microseconds of an overvoltage condition. It is a good idea to incorporate the crowbar directly into the transceiver. The main difference between the two circuits is that less complex circuit B depends on component tolerances for the exact trigger level, while the circuit A includes a unijunction trigger to permit precise setting of the operating point.

AC POWER-LINE CONNECTIONS MONITOR

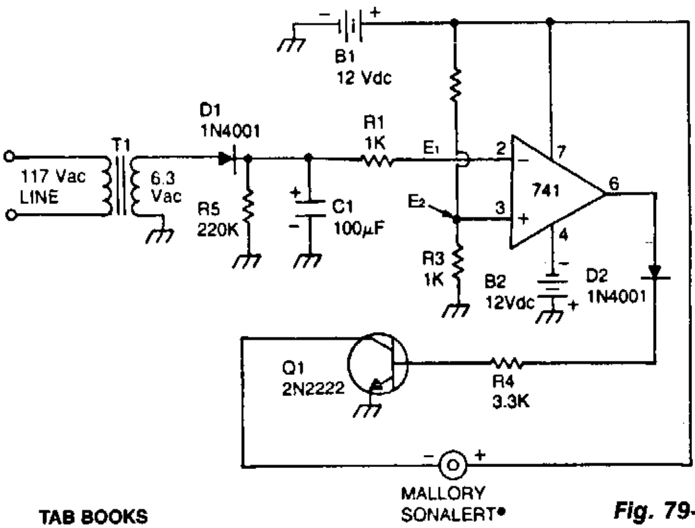


EDN

Fig. 79-2

A continuous glow signifies that everything is normal; a blinking or extinguished neon bulb indicates a broken earth-ground connection, or interchanged neutral and live wires.

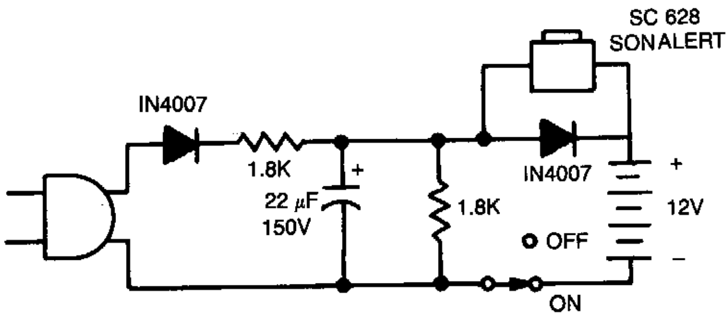
LINE-VOLTAGE MONITOR



TAB BOOKS *Fig. 79-3*

This circuit uses a type 741 op amp as a voltage comparator. One input of the 741 is connected to a reference voltage (a 12-V battery) through a resistor voltage divider. The potential at the noninverting input of the 741 is approximately 3 V. The inverting input of the op amp comparator is connected to the output of a line-operated 8-V power supply. When the ac power main fails, T1 will no longer be energized, so the charge stored in capacitor C1 will begin to discharge through resistor R5. When the capacitor voltage drops below the reference voltage of 3 V, the output of the comparator becomes high. This output condition will forward bias transistor Q1, causing the Sonalert to sound the alarm. The time constant of the R5/C1 combination is 22 seconds—long enough to prevent noise from triggering the alarm.

POWER-FAILURE ALARM



Reprinted with permission of William Sheets. *Fig. 79-4*

With power ac off, the alarm sounds when S1 is closed on. The 12-V battery is kept charged when the circuit is plugged in and the switch is left on.

AC CIRCUIT BREAKER

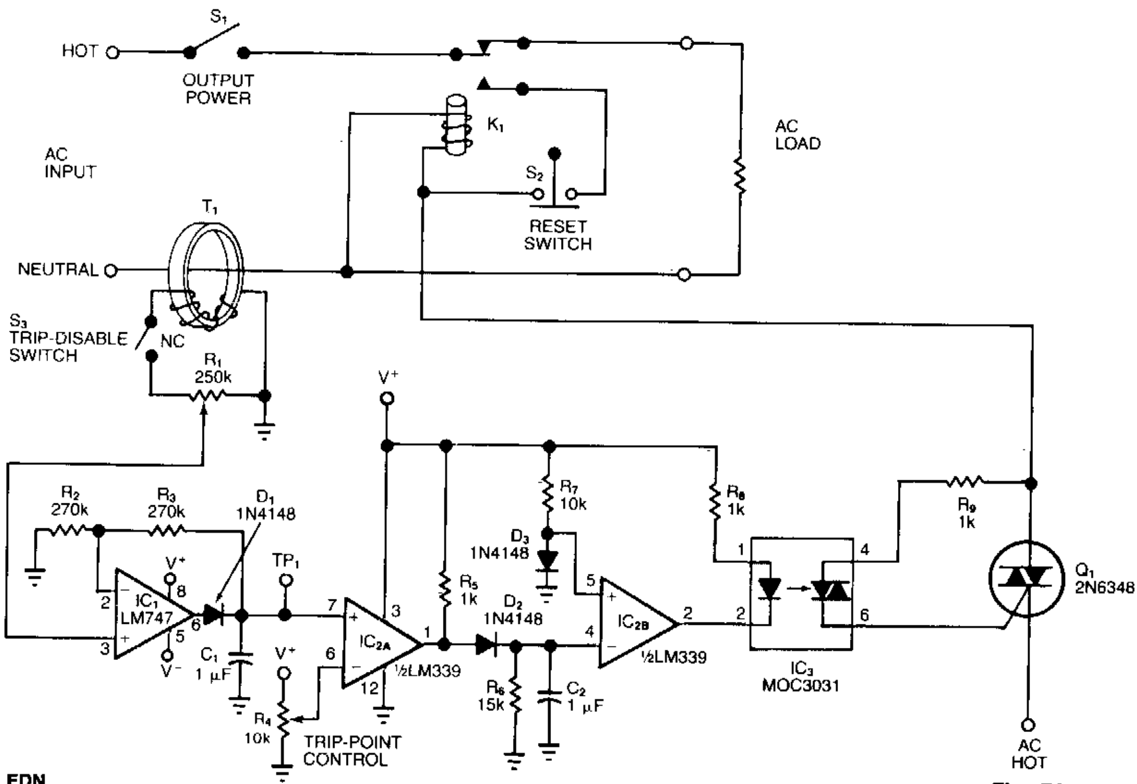


Fig. 79-5

The adjustable circuit breaker responds in 0.02 s under all conditions—provided you select a fast relay for K1. For moderate overload conditions, it's preferable to use the fuse or the fast-acting breaker. Toroid transformer T1 senses ac load current and produces an ac signal at the wiper of R1, when switch S3 is closed. Diode D1 rectifies this signal to produce a positive voltage at test point TP1. Because R1 allows you to calibrate this voltage, the circuit accommodates a variety of current-sense transformers. To calibrate the trip threshold, apply the maximum expected overload and adjust R1 until the TP1 voltage is 0.7 V below the positive saturation level for IC1. Then adjust R4 for the desired trip point. To reset the circuit breaker after it has tripped, open S1 or S2.

FAST OVERVOLTAGE PROTECTOR

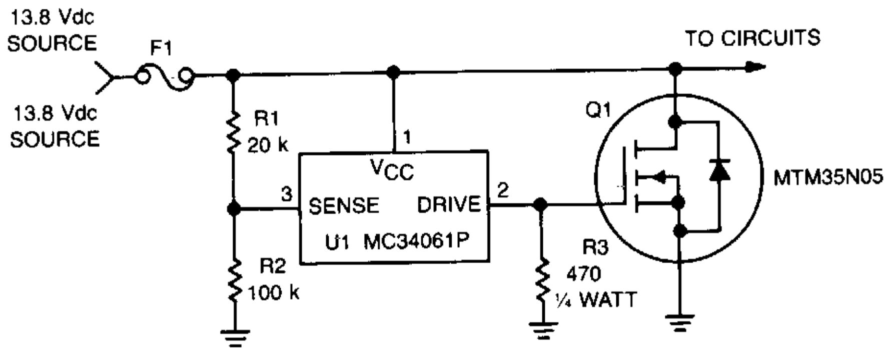


Fig. 79-6

This circuit protects expensive portable equipment against all types of improper hookups and environmental hazards that could cause an overvoltage condition. It operates very quickly and does not latchup, that is, it recovers when the overvoltage condition is removed. In contrast, SCR overvoltage circuits can latch and do not recover, unless the power is removed.

Here, U1 senses an overvoltage condition when the drop across R1 exceeds 2.5 V. This causes U1 to apply a positive signal to the gate of Q1, turning it on and shorting the line going to the external circuits. Fuse 1 opens if the transient condition lasts long enough to exceed the i^2t rating.

80

Proximity Sensors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

CAPACITIVE SENSOR ALARM

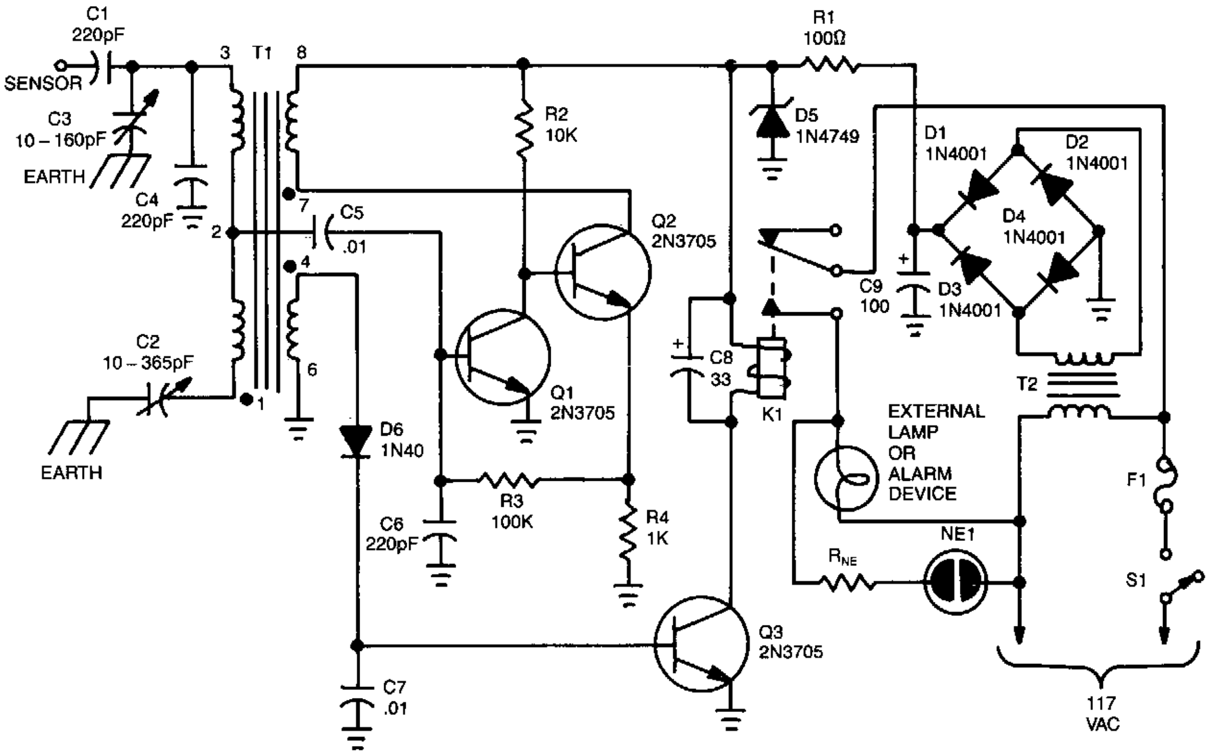


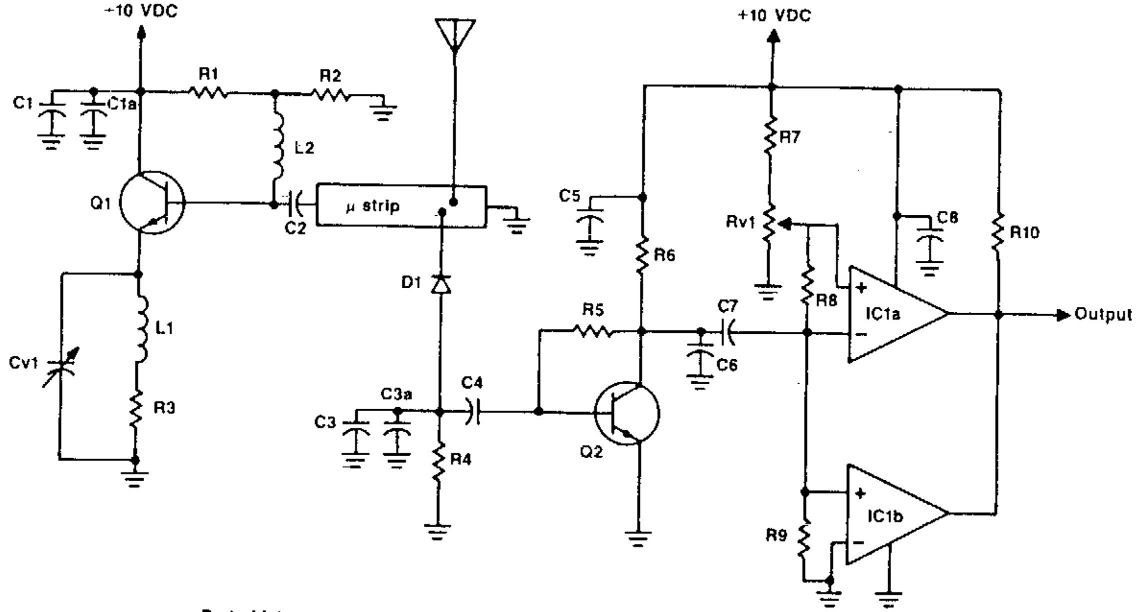
Fig. 80-1

The unit is constructed around a balanced-bridge circuit, using both capacitance and inductance. The bridge consists of capacitors C2 and C3, and the center-tapped winding of T1. One end of the bridge is coupled to ground by C4, while capacitance changes are introduced through C1. A small capacitance change unbalances the bridge and produces an ac signal at the base of Q1. Transistors Q1 and Q2 are connected to form a modified-Darlington amplifier. The collector load for Q2 is a separate winding of T1 that is connected out-of-phase with the incoming ac signal. That produces a large, distorted signal each time the bridge is unbalanced.

The distorted signal is taken from the bridge circuit by a third winding of transformer T1. That signal is then rectified by D6 and applied as a dc signal to the base of Q3. The applied signal energizes the relay, K1, as soon as the unbalanced condition occurs, and the relay drops out as soon as the circuit balance is restored. Of course, for normal alarm use, the relay should be made self-latching, so that the alarm condition remains in effect until the system is reset.

An audible alarm, such as a bell or klaxon horn, can be operated from the relay. If a silent alarm is needed, a light bulb can be used. Transformer T1 can be purchased as part #6182 from: Pulse Engineering, P.O. Box 12235, San Diego, CA 92112.

UHF MOVEMENT DETECTOR



Parts List

R1	3.9 k	C1	1 nF	D1	1SS97 or other Schottky type
R2, 4	1 k	C1a	470 uF	Q1	MRF961
R3	100	C2	47 nF	Q2	BC548
R5	2.2 M	C3a	1 uF	Cv1	2-7 pF miniature
R6	6.8 k	C4	22 uF	IC1	LM339 comparator
R7	100 k	C5, 8	100 uF	L1	5 turns 0.86 mm wire on 3.5 mm core
R8, 9, 10	22 k	C6	100 nF		
Rv1	1 k	C7	10 uF	L2	4 turns 0.86 mm wire on 3.5 mm core

D. Huisman, RF Design, December 1986, p. 41.

Fig. 80-2

The oscillator is a standard UHF design which delivers about 10 mW at 1.2 GHz. R1 and R2 bias the base of Q1 to 1.2 V via L2. Collector current is set by R3 to about 30 mA. C2 couples the base of Q1 to the stripline circuit. Tuning is provided by CV1, and C1 plus C1a decouple the collector. R2 and R3 are not decoupled, since this could cause instability.

Q2 is a simple one-transistor amplifier. C4 and C7 reduce gain below 1.5 and above 100 Hz; the remaining band of frequencies is amplified and passed on to the level detector. Two comparators of IC1 provide level detection. The trigger voltage is set by R7, Rv1, R8, and R9; it is adjustable from 8 to 60 mV by Rv1.

Positive voltage swings above the trigger level cause the IC1a output to become low, while negative swings cause IC1b to become low. C8 decouples IC1 from the power supply, and R10 is a pull-up resistor for the open collector output of IC1.

PROXIMITY SWITCH

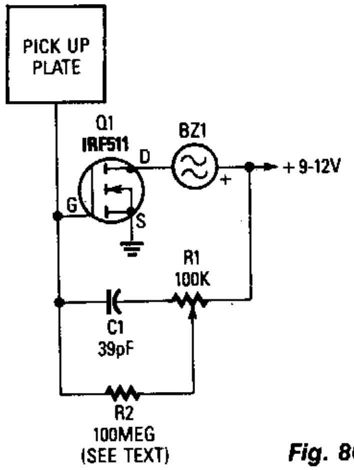


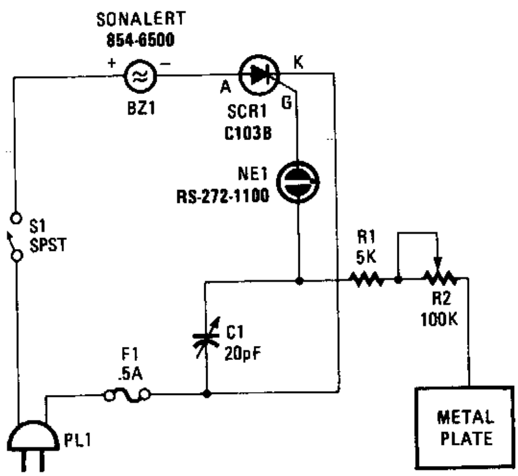
Fig. 80-3

POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

A 3-×-3-inch piece of circuit board, or similar size metal object which functions as the pick-up sensor, is connected to the gate of Q1. A 100-MΩ resistor, R2, isolates Q1's gate from R1, allowing the input impedance to remain very high. If a 100-MΩ resistor cannot be located, just tie five 22-MΩ resistors in series and use that combination for R2. In fact, R2 can be made even higher in value for added sensitivity.

Potentiometer R1 is adjusted to where the piezo buzzer just begins to sound off and then carefully backs off to where the sound ceases. Experimenting with the setting of R1 will help in obtaining the best sensitivity adjustment for the circuit. Resistor R1 can be set to where the pick-up must be contacted to set off the alarm sounder. A relay or other current-hungry component can take the place of the piezo sounder to control most any external circuit.

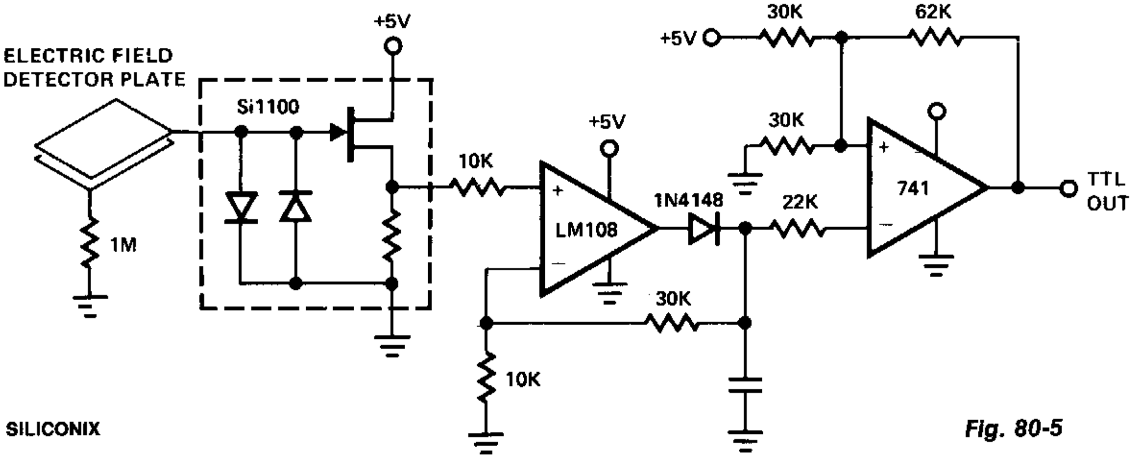
SCR PROXIMITY ALARM



HANDS-ON ELECTRONICS

Fig. 80-4

PROXIMITY SENSOR



SILICONIX

Fig. 80-5

The Si1100 series circuit input is connected to a capacitive field sensor—possibly a piece of double-sided circuit board. Any induced voltage change on the plate is fed to the input of the peak detector section of the op-amp circuit. The Schmitt trigger monitors the voltage across the capacitor and changes its output state when the capacitor voltage crossed the 2.5-trigger point. The output from the Schmitt trigger switches between 0 and 5 V and is microprocessor compatible for sensor applications, such as computer-controlled intruder alarms.

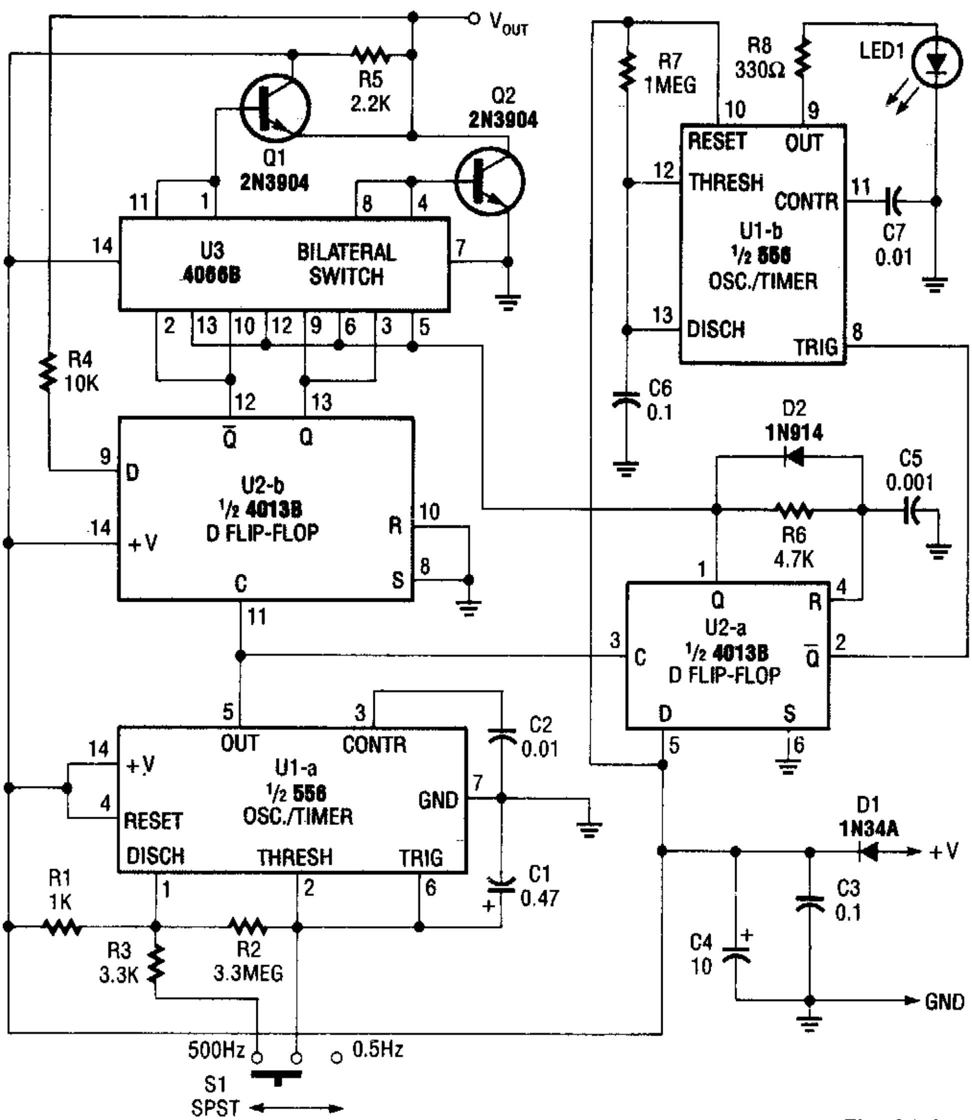
82

Ramp Generators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Logic Pulser
300-V Pulse Generator
Very Low Duty-Cycle Pulse Generator
Wide-Ranging Pulser
CMOS Short-Pulse Generator
Voltage Controller/Pulse-Width Generator

LOGIC PULSER



POPULAR ELECTRONICS

Fig. 81-1

The pulser generates pulses at a user-selected frequency of 0.5 or 500 Hz, with a pulse width of about 5 ms. If the input to be pulsed is already being driven high or low by another output, the pulser automatically pulses the input to the opposite logic state. The pulser is powered by the circuit under test, and operates from supplies of from +5 to +15 Vdc.

300-V PULSE GENERATOR

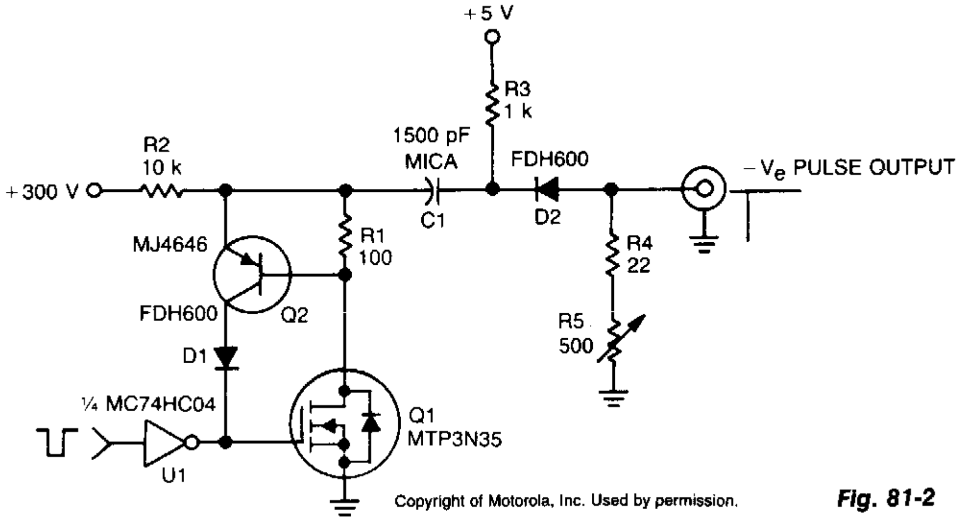
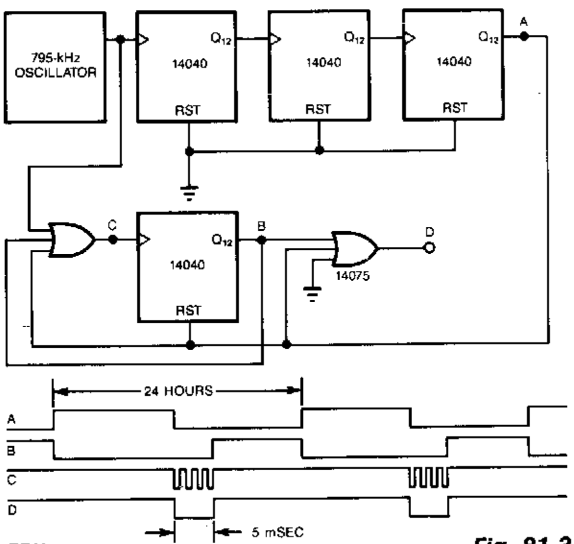


Fig. 81-2

In this TMOS pulser, a negative-going pulse is applied to U1, a high-speed CMOS buffer, which directly drives the gate of Q1, an MTP3N35. If only a 100-V pulse is required, the MTA6N10 can be used. The pulse output across R2 is differentiated by R3/C1 and appears as a negative-going spike at the output terminal.

VERY LOW DUTY-CYCLE PULSE GENERATOR

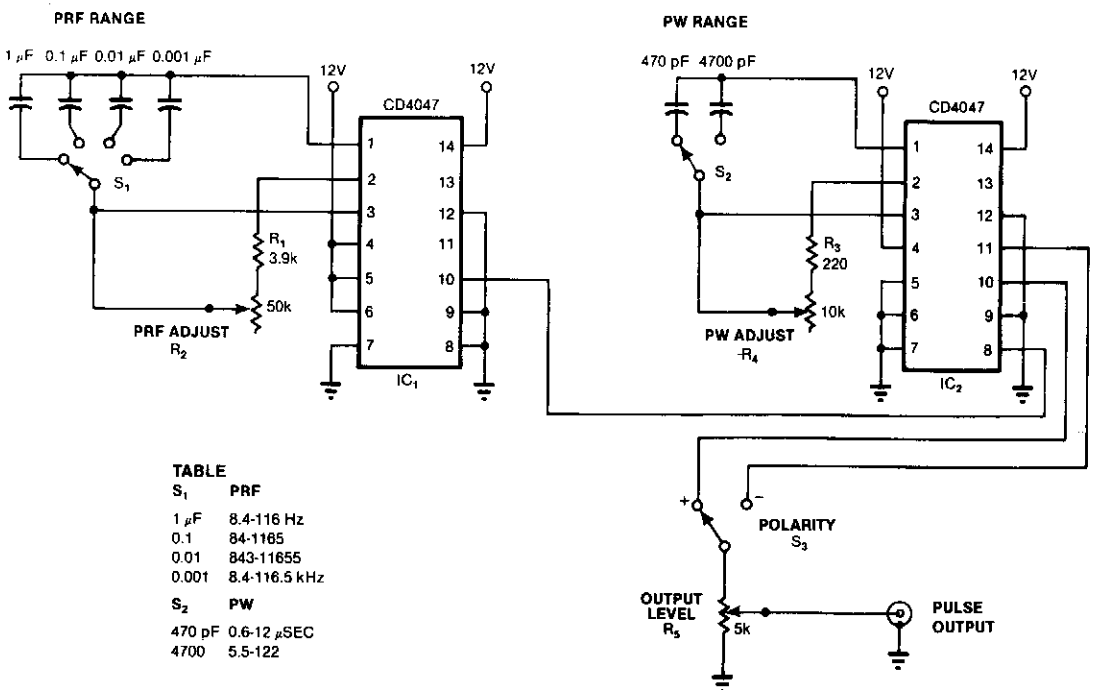


EDN

Fig. 81-3

Using a precision oscillator and a few CMOS counters, you can build a precise, very low duty-cycle pulse generator. You can add as many counters as you desire to make the period as long as you wish. The circuit will generate a pulse about 5 ms long every 24 hours—one 5.9 −6% duty cycle.

WIDE-RANGING PULSER



TABLE

S_1	PRF
1 μF	8.4-116 Hz
0.1	84-1165
0.01	843-11655
0.001	8.4-116.5 kHz

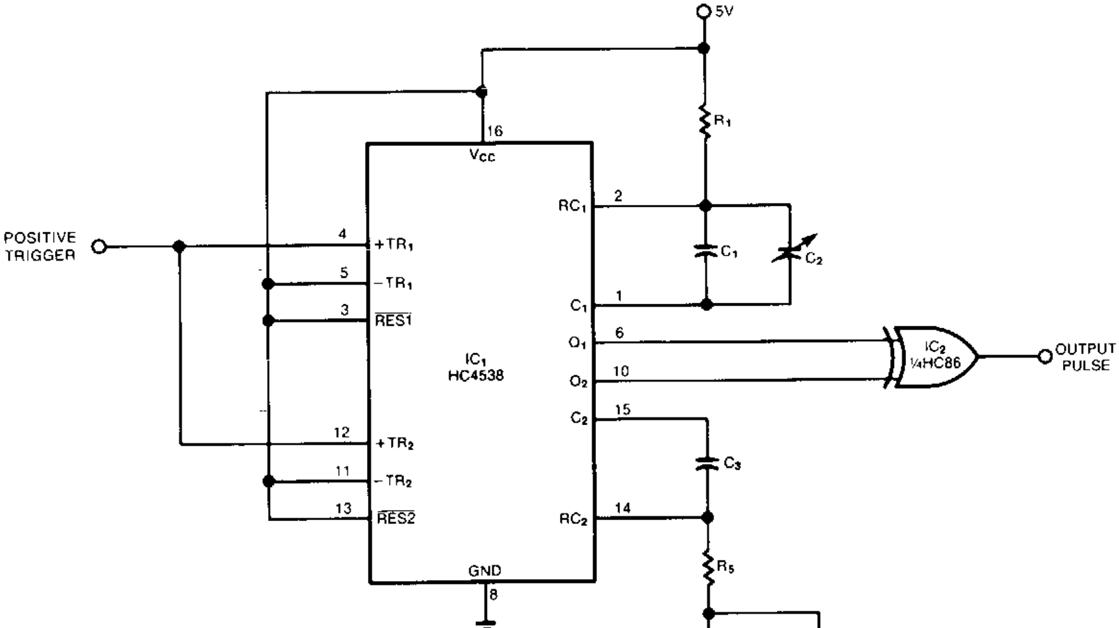
S_2	PW
470 pF	0.6-12 μSEC
4700	5.5-122

EDN

Fig. 81-4

An output pulse's characteristics depend upon two multivibrator's timing components. IC1's free-running astable-mode frequency sets the pulse's prf, whereas the pulse's width comes from IC2's monostable operation.

CMOS SHORT - PULSE GENERATOR



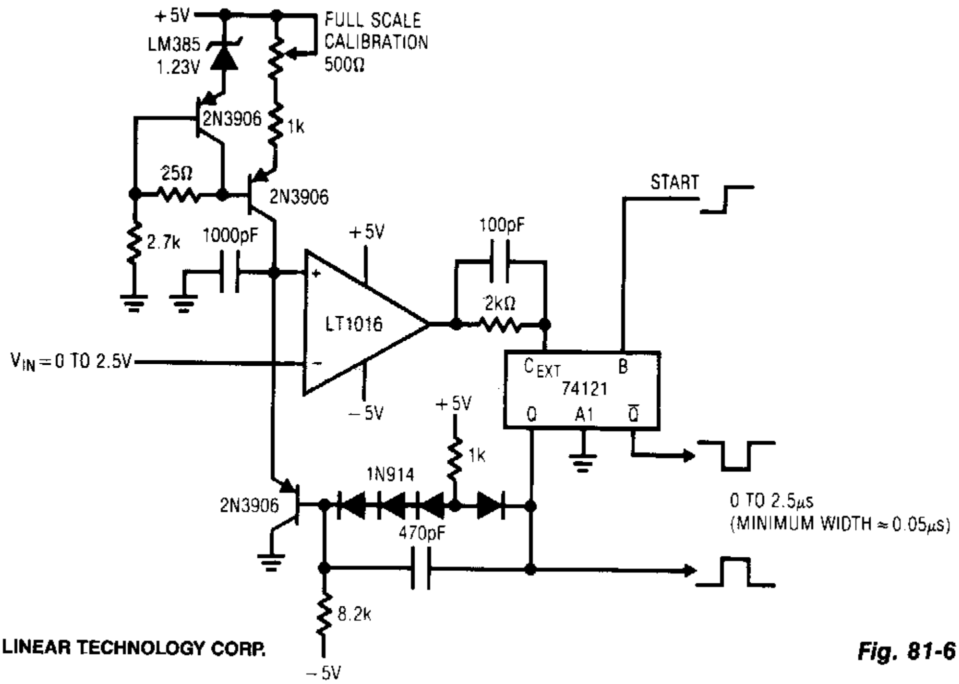
NOTE:
FOR A 50- TO 250-nSEC PULSE OUTPUT:
R_1 – R_5 – 10k
R_3 = 10k, 10T
R_4 – 4.7k
C_1 = C_3 = 100 pF, SILVER-MICA OR POLYSTYRENE
C_2 = 5- TO 35-pF CERAMIC TRIMMER CAPACITOR

EDN

Fig. 81-5

Comprising two low-power, CMOS chips, the pulse generator produces a precise pulse width in the 50 to 500 ns range. IC1 is a dual monostable multivibrator (one shot) in which each positive trigger pulse initiates simultaneous positive output pulses at pins 6 and 10. In response, XOR gate IC2 produces a positive pulse whose duration is equal to the difference between the two input-pulse durations. Section 1 of the one shot generates an approximate 1-μs reference pulse—shorter pulses are more susceptible to manufacturing variations caused by parasitic layout capacitance. Variable capacitor C2 lets you adjust this pulse width. Section 2 of the one shot generates a variable-length pulse; you adjust its width by using potentiometer R3. Resistors R4 and R5 set the output pulse's maximum and minimum width, respectively. Because the XOR gate's rise and fall times are about 20 ns for reasonable values of load capacitance, you should calibrate the circuit using C2 for a minimum output-width of 50 ns.

VOLTAGE CONTROLLER/PULSE-WIDTH GENERATOR



LINEAR TECHNOLOGY CORP.

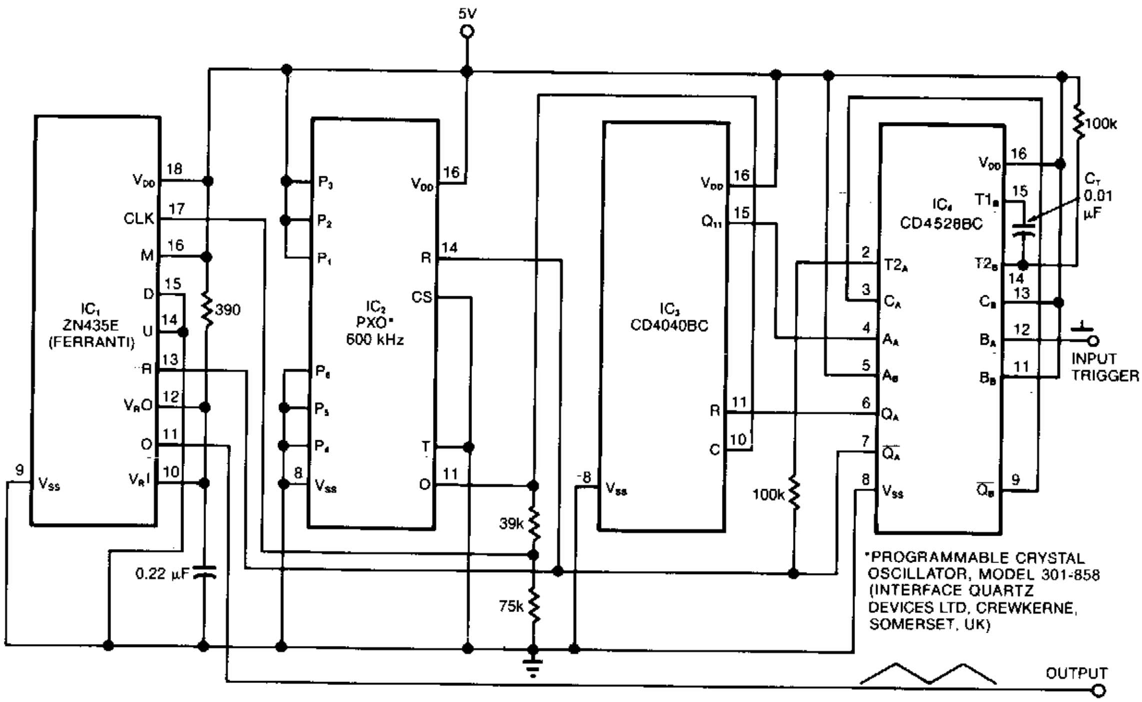
Fig. 81-6

82

Ramp Generators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

ACCURATE RAMP GENERATOR



EDN

Fig. 82-1

The ramp generator, an inexpensive alternative to commercial function generators, provides a more linear and repeatable output than conventional analog integrators. The circuit provides a triangle waveform in burst mode; in this case, two cycles of 10.24 ms each per input trigger pulse. IC4 is a dual monostable multivibrator (one shot) in which the A side is configured as a latch (see *Multivibrator IC performs extra tasks*, EDN, September 6, 1984, p. 232). The rising edge of each input pulse triggers the B side, producing at pin 9 an output pulse whose duration depends on the timing capacitor's, C_T, value—A 0.01-μF value gives a 500 μs pulse. This output provides a reset to the A side latch. While the latch is reset with Q_A high, Q_A low, the other three ICs are active. The P1 through P6 connections, as shown, set oscillator IC2's frequency to 50 kHz at pin 11.

Counter IC3 counts upward. The output at pin 11 of multifunction converter IC1 ramps up to full-scale, reverses, ramps down to zero, and then repeats this sequence of events. As this output completes its second cycle, IC3 reaches a count of 1024, causing the Q11 output to become high and toggle the IC4 latch. The resulting change of state on Q_A and Q_A resets the other three ICs, terminating further activity until the arrival of the next input trigger pulse. IC2 is included for its synchronous-reset capability, and it therefore drives the internal clock of IC1, which cannot be synchronously reset. Still IC2 can be omitted in some applications. The circuit operates from a 5-V supply. You can modify the output by changing IC2's frequency and IC3's output connection.

INTEGRATOR/RAMP GENERATOR WITH INITIAL CONDITION RESET

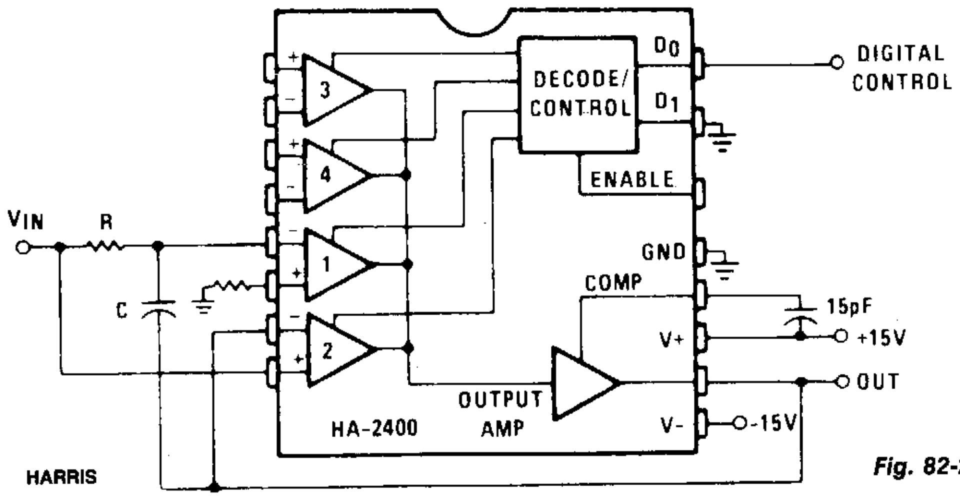


Fig. 82-2

Channel 1 is wired as a conventional integrator, and channel 2 as a voltage follower. When channel 2 is switched on, the output will follow V_{IN} and C will discharge to maintain 0 V across it. When channel 1 is then switched on, the output will initially be at the instantaneous value of V_{IN}, and then will commence integrating towards the opposite polarity. This circuit is particularly suitable for timing ramp generation using a fixed dc input. Many variations, such as building programmable time constant integrators, are possible.

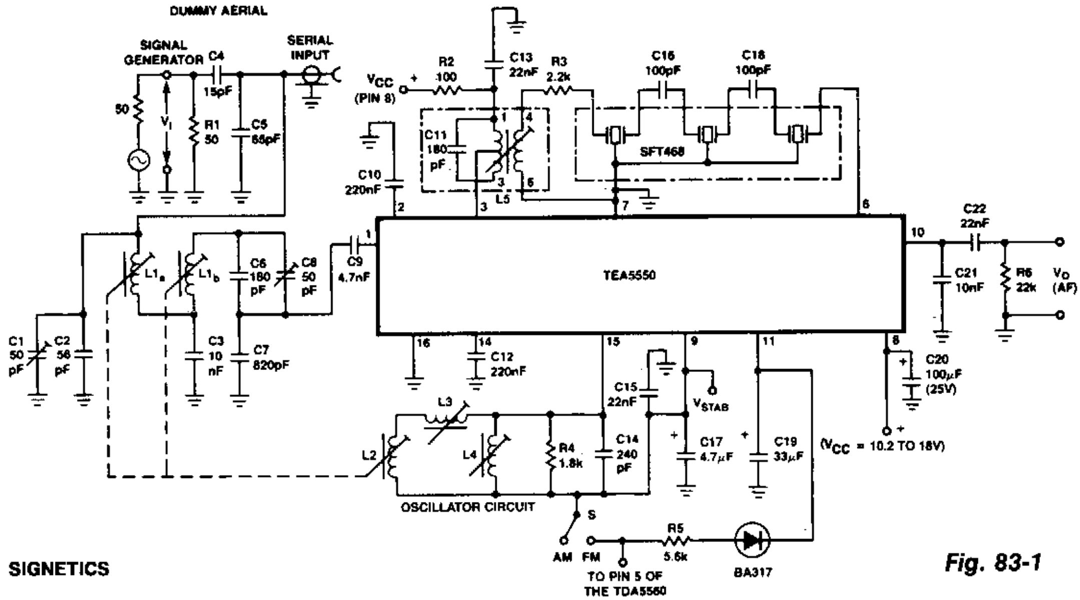
83

Receivers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

- AM Radio
- FM Tuner
- FM MPX/SCA Receiver
- Narrow-Band FM Receiver
- Low-Cost Line Receiver
- FSK Data Receiver
- Simple Ham-Band Receiver
- Digital Data Line Receiver
- Integrated AM Receiver

AM RADIO



SIGNETICS *Fig. 83-1*

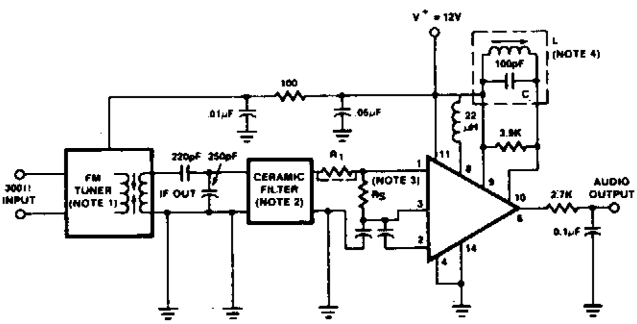
NOTES:

Coil data: L1a, L1b, L2 = Tuning coils: ALPS unit MMK IIEII (for coil connections see Figure 7)
L3 = Trimming coil (4.7µH); catalog number 3122 138 27460
L4 = Padding coil (200µH); catalog number 3111 118 23510
L5 = IF coil; catalog number 3122 138 91481

S is AM/FM switch; for printed circuit board see Figures 7 and 8.

This circuit diagram is for a double-tuned, AM-channel, in-car radio receiver using the TEA5550.

FM TUNER



NOTES:
All resistor values are typical and in ohms.
1. Waller 4SN3FIC or equivalent.
2. Murata SFG 10.7mA or equivalent.
3. R_S will affect stability depending on circuit layout. To increase stability R_S is decreased. Range of R_S is 330
4. L tunes with 100pF (C) at 10.7MHz Q_O unloaded ≃ 75 (G.I. EX27825 or equivalent).

Performance data at f_O = 98MHz, f_{MOD} = 400Hz, deviation = ± 74kHz.
± 74kHz.

−3dB limiting sensitivity	2µV (antenna level)
20dB quieting sensitivity	1µV (antenna level)
30dB quieting sensitivity	1.5µV (antenna level)

SIGNETICS *Fig. 83-2*

FM MPX/SCA RECEIVER

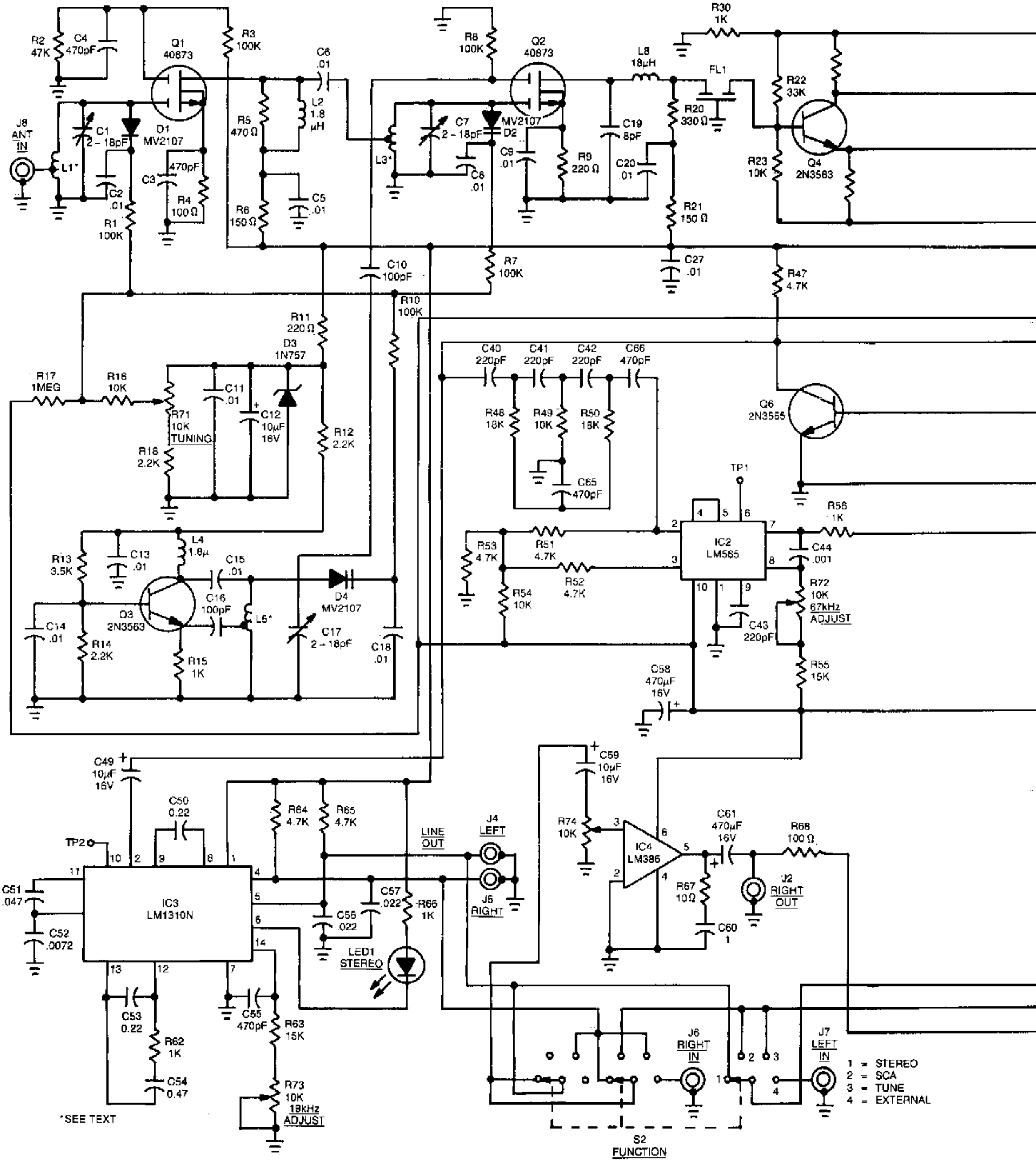
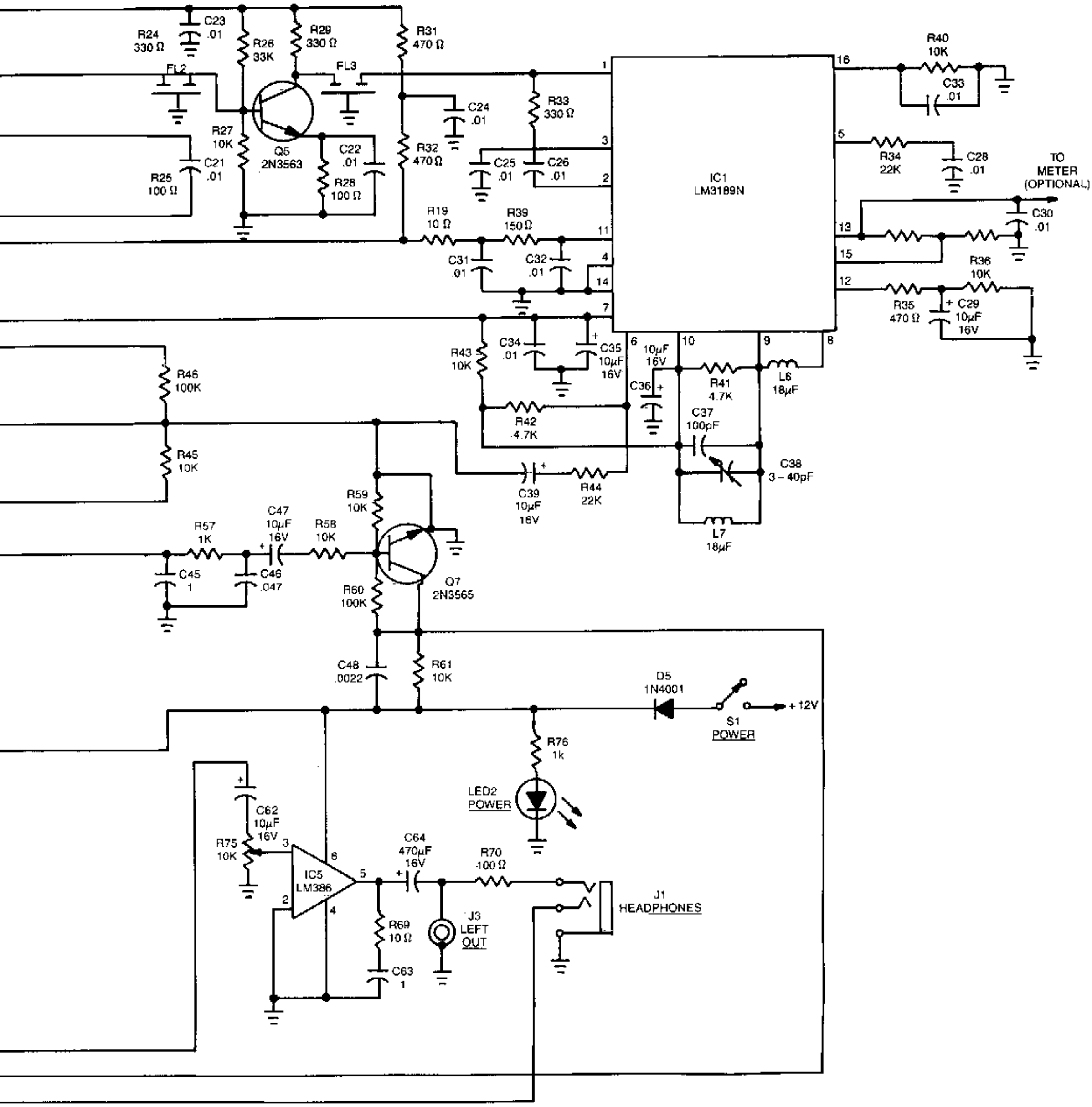


Fig. 83-3



This receiver is capable of better than 1.5 μ VIHF sensitivity and uses MOSFET front-end circuitry with varactors to eliminate conventional bulky tuning capacitors. It also features high dynamic range, ceramic i-f filters requiring no alignment, and a quadrature-type detector with excellent limiting and AM rejection capability. The receiver operates from nominal 12-V supply. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

NARROW-BAND FM RECEIVER

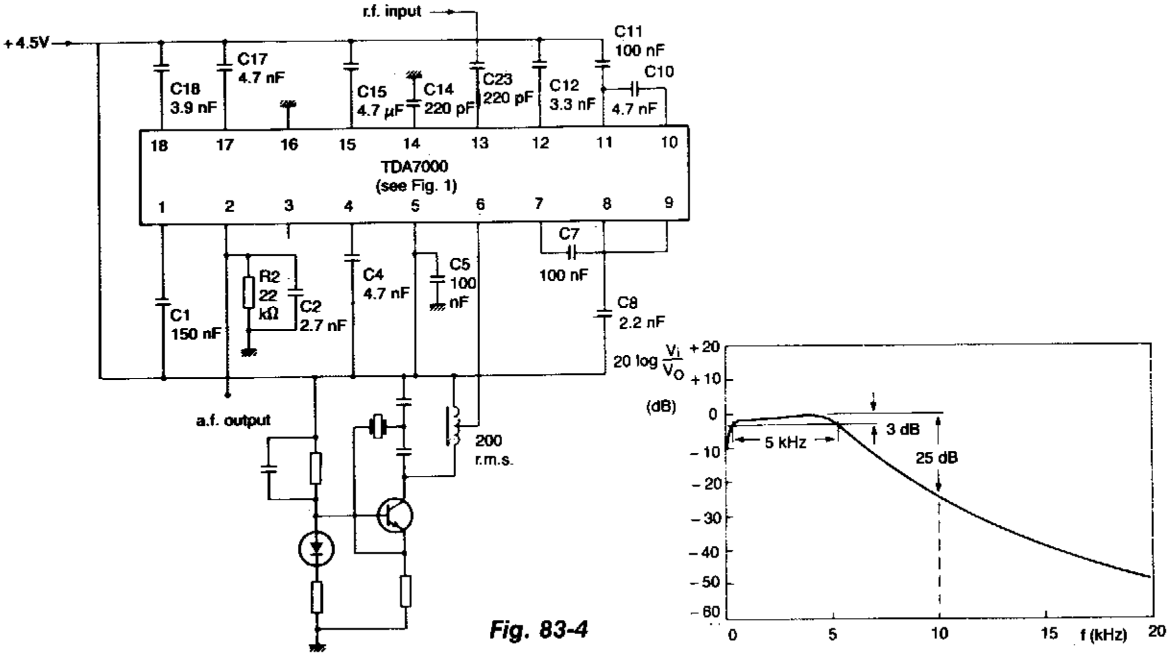
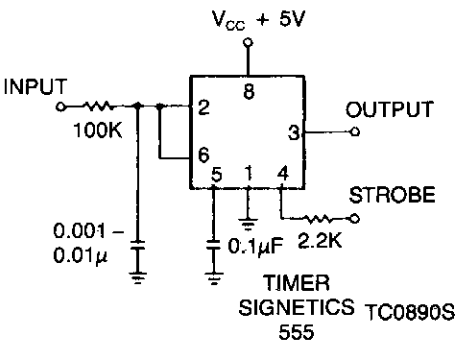


Fig. 83-4

The local oscillator is crystal-controlled and the i-f swing is hardly compressed. The deviation of the transmitted carrier frequency, because of modulation, must therefore be limited to prevent severe distortion of the demodulated audio signal. The component values result in an i-f of 4.5 kHz and an i-f bandwidth of 5 kHz. If the i-f is multiplied by N, the values of capacitors C17 and C18 in the all-pass filters, and the values of filter capacitors C7, C8, C10, C11, and C12 must be multiplied by $1/N$. For improved i-f selectivity to achieve greater adjacent channel attenuation, second-order networks can be used in place of C10 and C11.

LOW-COST LINE RECEIVER



NOTE:
ALL RESISTOR VALUES ARE IN OHMS.

SIGNETICS

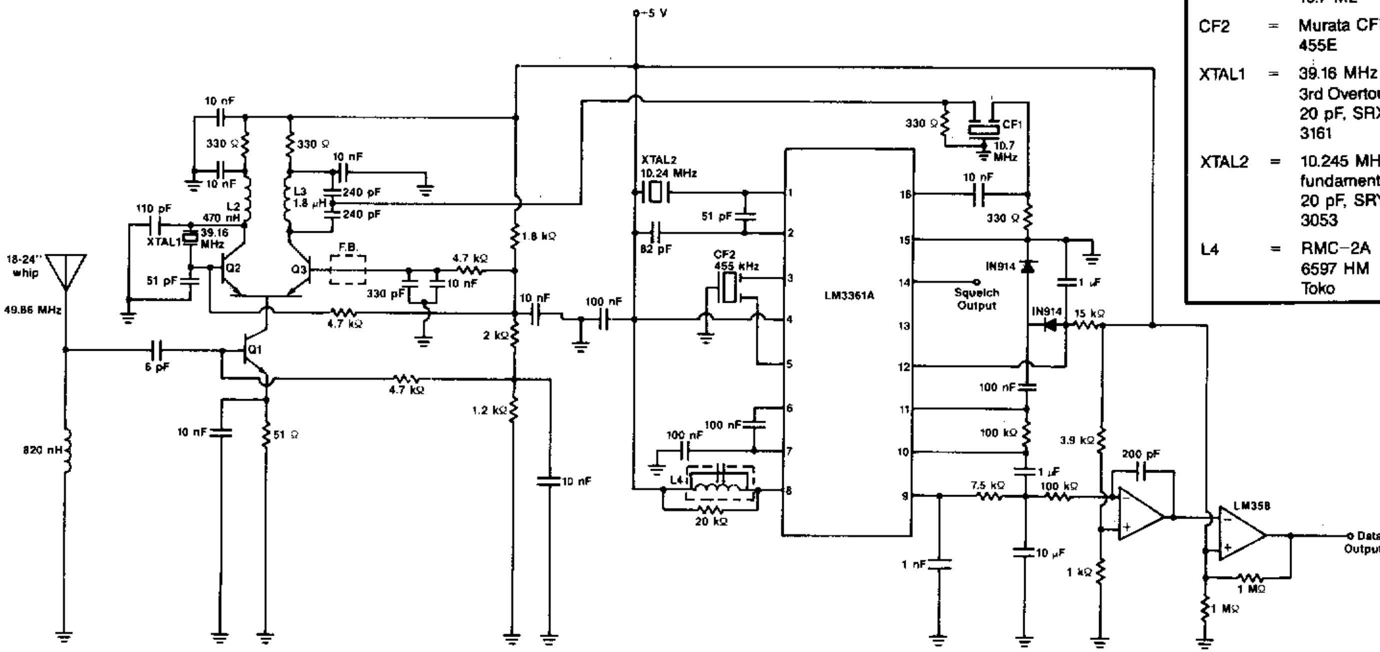
Fig. 83-5

This timer makes an excellent line receiver for control applications involving relatively slow electromechanical devices. It can work without special drivers over single, unshielded lines.

FSK DATA RECEIVER

Parts List

$Q_{1,2,3}$	=	MPSH 20
CF1	=	Murata SFE 10.7 ML
CF2	=	Murata CFU 455E
XTAL1	=	39.16 MHz 3rd Overtone, 20 pF, SRX 3161
XTAL2	=	10.245 MHz fundamental, 20 pF, SRY 3053
L4	=	RMC–2A 6597 HM Toko

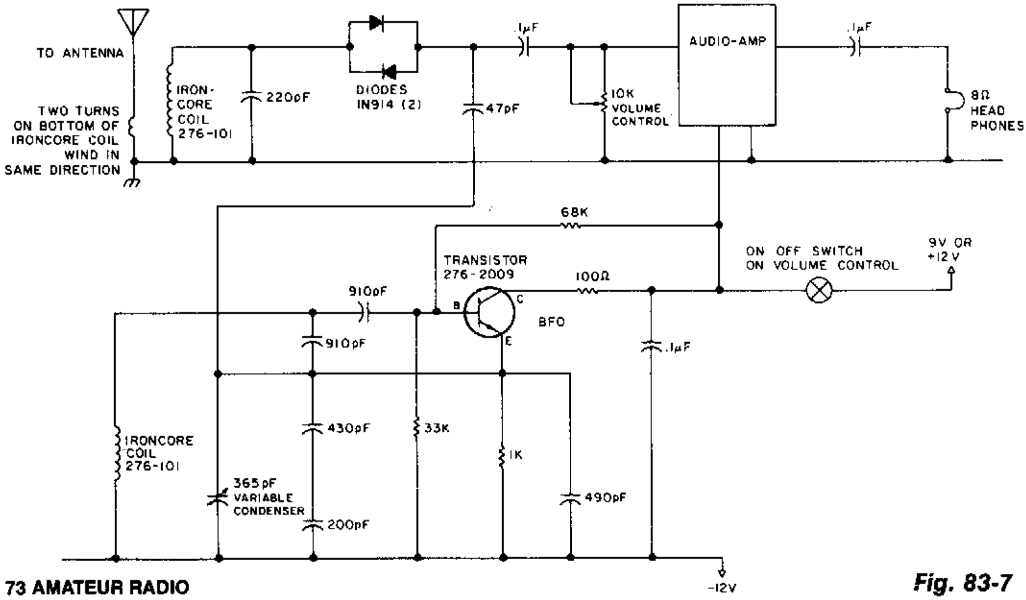


M. Lee, "A Simple FSK Data Receiver," RF Design, March 1987, p. 46.

Fig. 83-6

The various signal frequencies are obtained for an incoming carrier centered at 49.86 MHz. The receiver employs double conversion, with i-fs at 10.7 MHz and 455 kHz. Ceramic filters are used in both i-fs for selectivity and reduced-coil count. A quadrature detector is used to recover the baseband data, and an integrator and Schmitt trigger are used to filter the demodulated output. Also included is a squelch circuit that functions as a status line, and the open-collector output switches high when a signal is received. The LM3361A functions as the 2nd LO, 2nd mixer, limiting i-f, quadrature detector, and squelch; yet, it consumes less than 4 mA from a 5-V logic supply. The entire receiver requires approximately 10 mA.

SIMPLE HAM-BAND RECEIVER

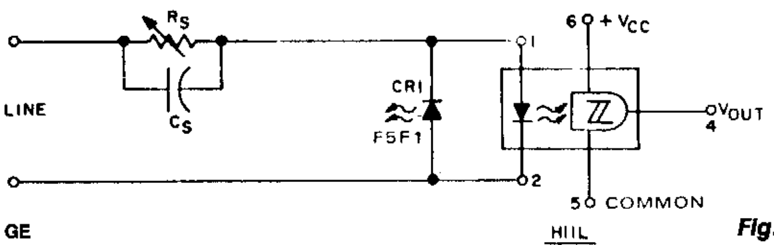


73 AMATEUR RADIO

Fig. 83-7

This circuit is configured for the 80m band. The 365-pF, broadcast-band variable capacitor should have a vernier drive with a six-to-one ratio, which makes tuning easier by separating the stations on the dial. A good antenna and ground are also recommended. The Radio Shack iron-core chokes (276-101), used in the bfo part of the circuit, can be calibrated by listening for the bfo signal in a calibrated receiver.

DIGITAL DATA LINE RECEIVER

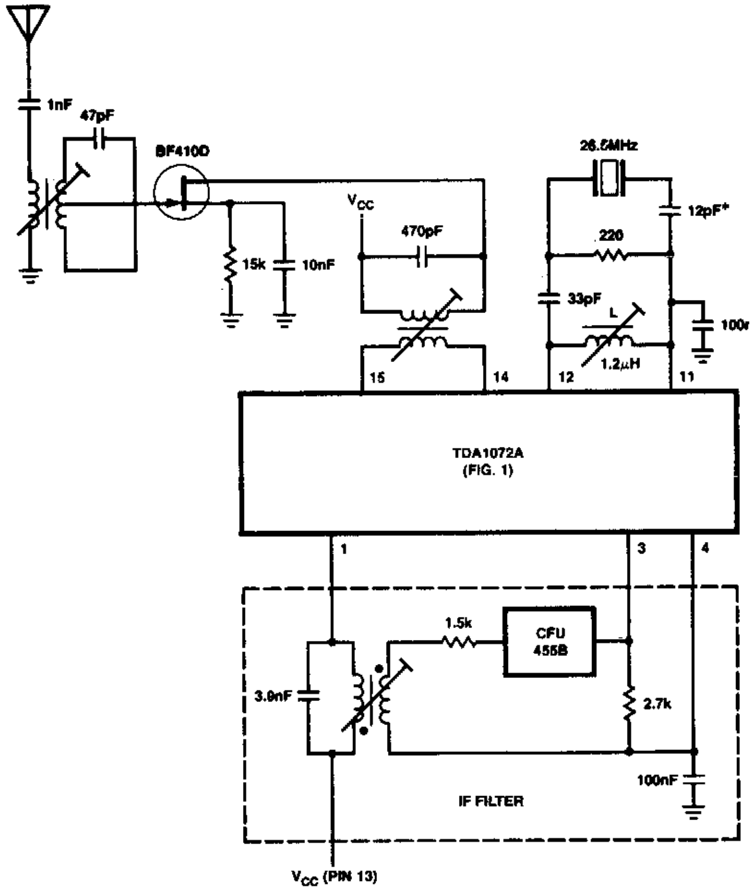


GE

Fig. 83-8

When digital data is transmitted over long lines (longer than 1 meter), proper transfer is often disturbed by the parasitic effects of ground level shifts and ground loops, as well as by extraneous noise picked up along the way. An optocoupler, such as the H11L, combining galvanic isolation to minimize ground loop currents and their concomitant common-mode voltages, with predictable switching levels to enhance noise immunity, can significantly reduce erratic behavior. Resistor R_S is programmed for the desired switching threshold, C_S is an optional speed-up capacitor, and CR1 is an LED used as a simple diode to provide perfect line balance and a discharge path for C_S if the speed-up capacitor is used.

INTEGRATED AM RECEIVER



NOTE:
A crystal Oscillator is used so that a narrow-band hybrid IF filter can be used.

SIGNETICS

Fig. 83-9

This circuit has aerial and local oscillator circuits for a 27-MHz receiver for remote control of garage doors, projectors, curtains, etc.

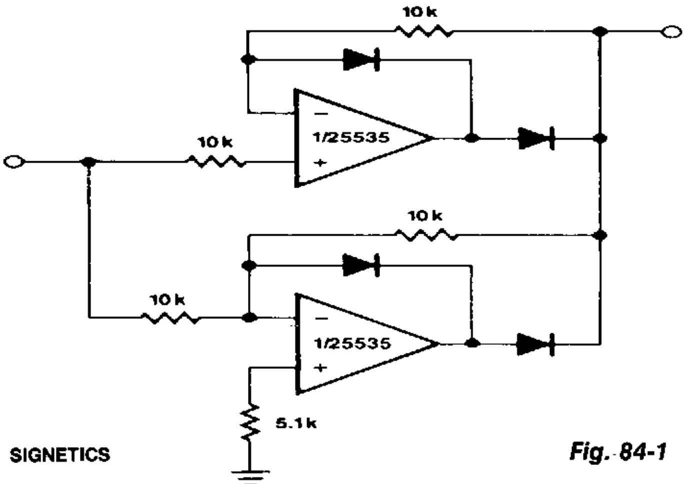
84

Rectifier Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Precision Full-Wave Rectifier
Diodeless Precision Rectifier

PRECISION FULL-WAVE RECTIFIER

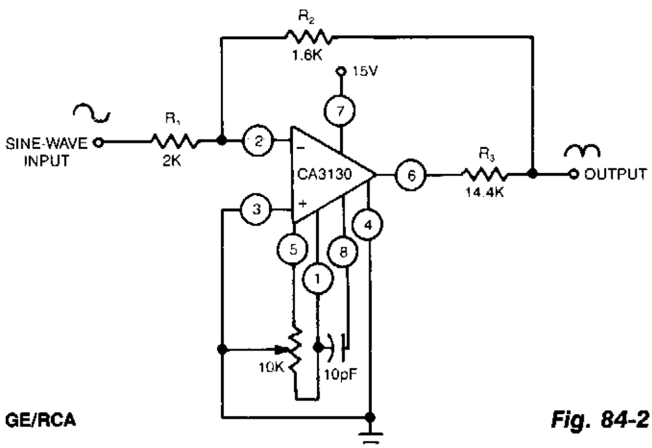


SIGNETICS

Fig. 84-1

This circuit provides accurate full-wave rectification. The output impedance is low for both input polarities, and the errors are small at all signal levels. Note that the output will not sink heavy currents, except a small amount through the 10-KΩ resistors. Therefore, the load applied should be referenced to ground or a negative voltage. The reversal of all diode polarities will reverse the polarity of the output. Since the outputs of the amplifiers must slew through two diode drops when the input polarity changes, the 741-type devices give 5% distortion at about 300 Hz.

DIODELESS PRECISION RECTIFIER



GE/RCA

Fig. 84-2

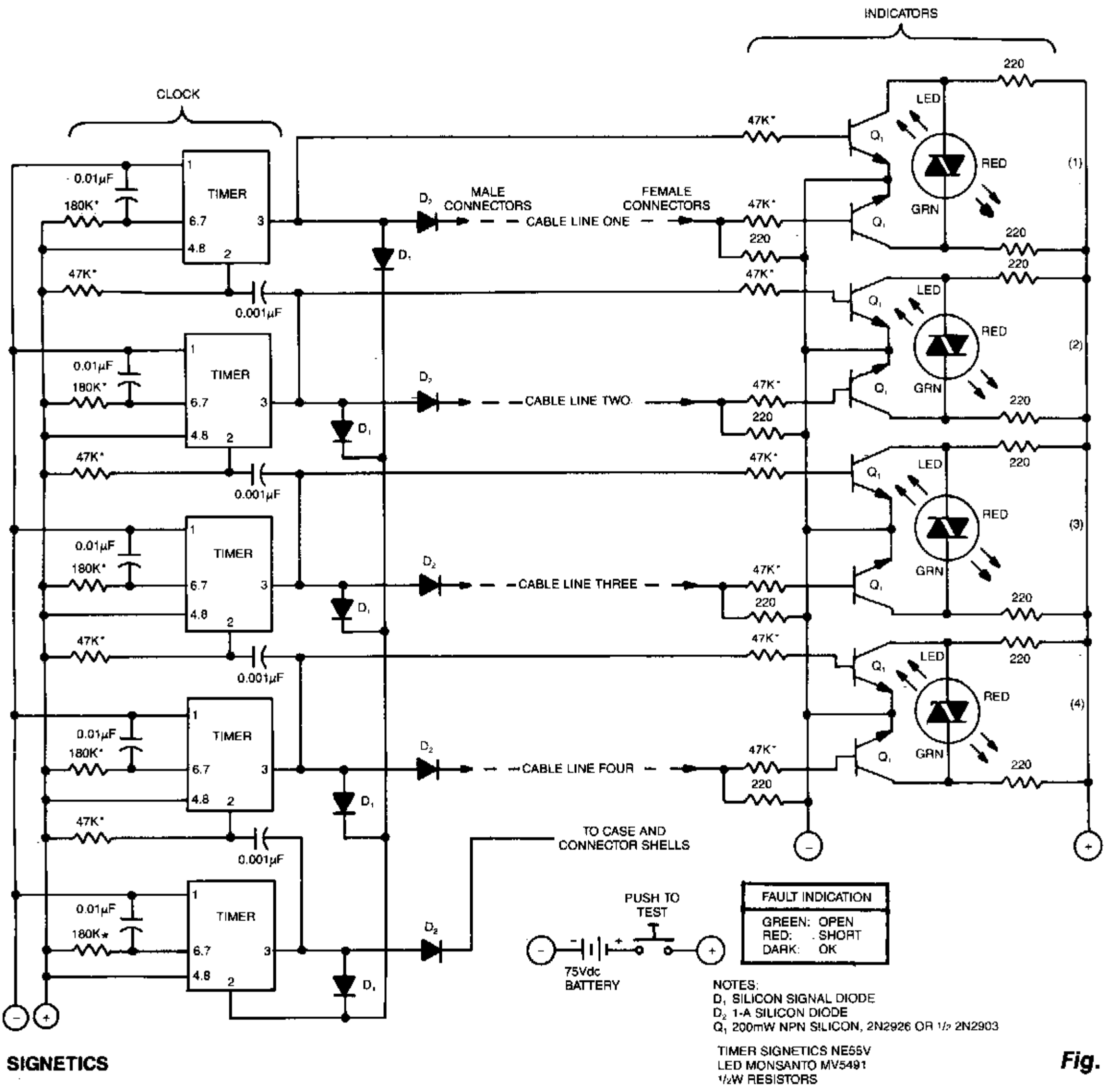
A CA3130 BiMOS op amp, acts as an attenuator for positive inputs and as a conventional op amp for negative signals. With 1-V rms input and a circuit gain of 0.8, its frequency response is −1% at 60 kHz and −1 dB at 300 kHz.

85

Resistance/Continuity Meters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

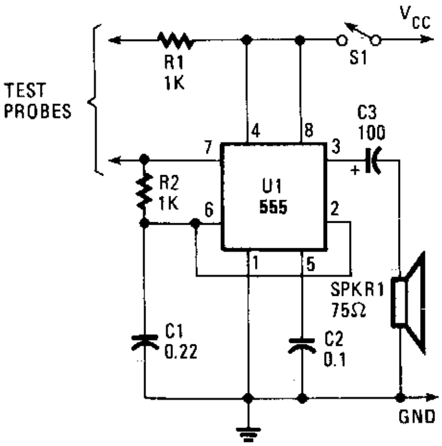
CABLE TESTER



SIGNETICS ***Fig. 85-1***

This compact tester checks cables for open-circuit or short-circuit conditions. A differential transistor pair at one end of each cable line remains balanced as long as the same clock pulse generated by timer IC appears at both ends of the line. A clock pulse, just at the clock end of the line, lights a green LED, and a clock pulse, only at the other end, lights a red LED.

CONTINUITY TESTER

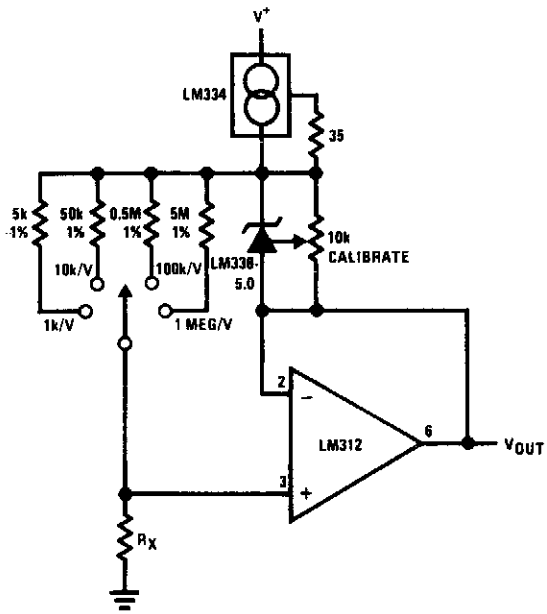


HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 85-2

The continuity tester feeds a voltage through the positive probe to the circuit-under-test, while the negative probe serves as the return line. Voltage that returns to the tester through the negative probe triggers the circuit, giving an audible indication of continuity.

LINEAR OHMMETER



NATIONAL SEMICONDUCTOR CORP.

Fig. 85-3

86

Rf Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

1296-MHz SOLID-STATE POWER AMPLIFIER

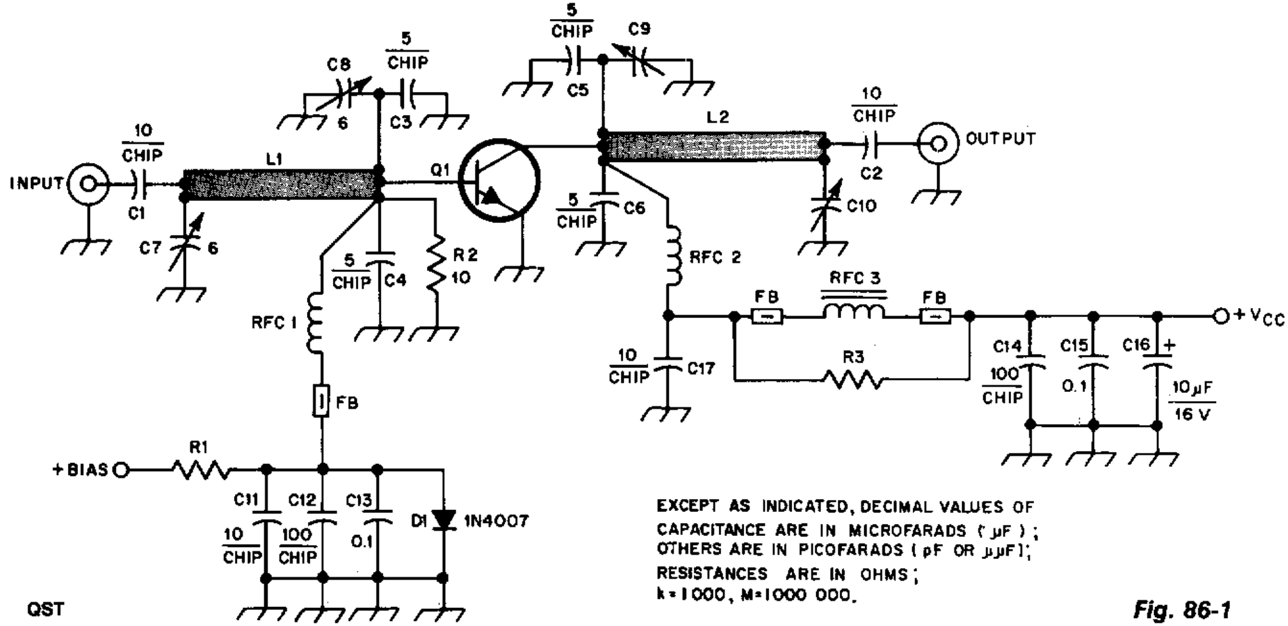


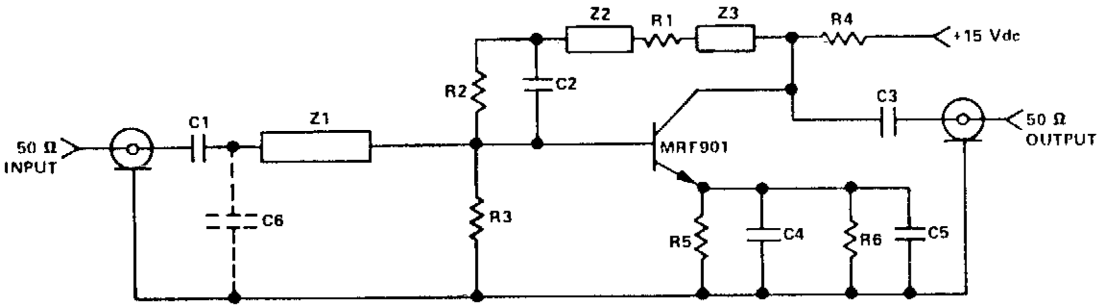
Fig. 86-1

Fig. 1—Schematic diagram of the NEL1306 and NEL1320 1296-MHz solid-state power amplifiers. The schematic is identical for both versions. Component values are the same except as noted.

C1, C2, C11, C17—10-pF chip capacitor.
C3, C4, C5, C6—3.6- to 5.0-pF chip capacitor.
C7, C8—1.8- to 6.0-pF miniature trimmer capacitor (Mouser 24AA070 or equiv. See text).
C9, C10—Same as C7 and C8 for the NEL1306 amplifier. For the NEL1320 version, 0.8- to 10-pF piston trimmers are used (Johanson 5200 series or equiv.).
C12, C14—100-pF chip capacitor.
C13, C15—0.1-µF disc ceramic capacitor.
C16—10-µF electrolytic capacitor.
D1—1N4007 diode.
L1, L2—30-ohm microstripline, ¼-wavelength long (see text).
Q1—NEC NEL130681-12 (6 W) or NEL132081-12 (18 W) transistor.
R1—82- to 100-Ω resistor, 2-W minimum. Vary for specified idling current.
R2—10-Ω, ¼-W carbon-composition resistor with "zero" lead length. See text.
R3—15-Ω, 1-W carbon-composition resistor.
RFC1—3t no. 24 wire, 0.125 inch ID, spaced 1 wire diam.
RFC2—1t no. 24 wire, 0.125 inch ID, spaced 1 wire diam.
RFC3—1-µH RF choke; 18t no. 24 enam. close-spaced on a T50-10 toroid core.

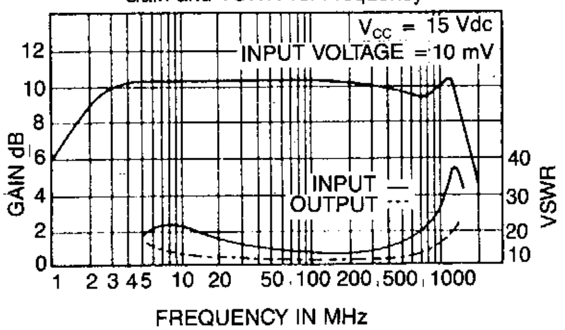
The design incorporates 30-Ω, $1/4\lambda$ microstrip lines on the input and output. C3, C4, C7, and C8, along with L1, form a pi network that matches the low-input impedance of the device to 50 Ω. C5, C6, C9, C10, and 30-Ω transmission line L2 form an output pi network that maximizes power transfer to 50 Ω. C10 is not always necessary, depending on variations among devices and circuit-board material. Bias is provided by R1, R2, and D1. R1 can be optimized, if desired, to adjust the collector idling current.

10 dB-GAIN AMPLIFIER



C1-C3	– 2200 pF chip capacitor	Z3	– 0.3″ × 0.125″ microstrip line	R4	– 560Ω carbon resistor
C4, C5	– 6.5 pF chip capacitor	R1	– 200Ω 1/8″ W, ±5% carbon resistor	R5, R6	– 15Ω ±5% chip resistor
C6	– Optional 2.1 pF chip capacitor	R2	– 4.3kΩ carbon resistor	Substrate	– 1 oz. copper, double-sided glass Teflon® board 0.0625″ thick, ε ≈ 2.5
Z1	– 0.3″ × 0.125″ microstrip line	R3	– 680Ω carbon resistor		
Z2	– 0.15″ × 0.125″ microstrip line				

Gain and VSWR vs. Frequency



Gain and VSWR vs. Frequency with Matching Capacitor C6

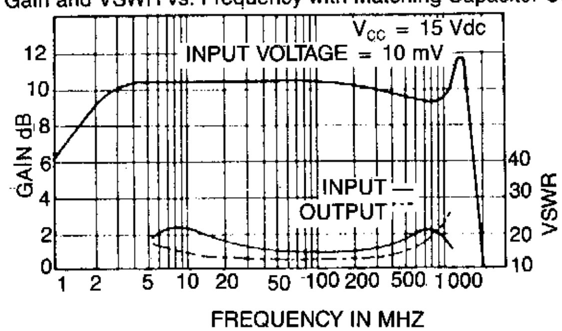
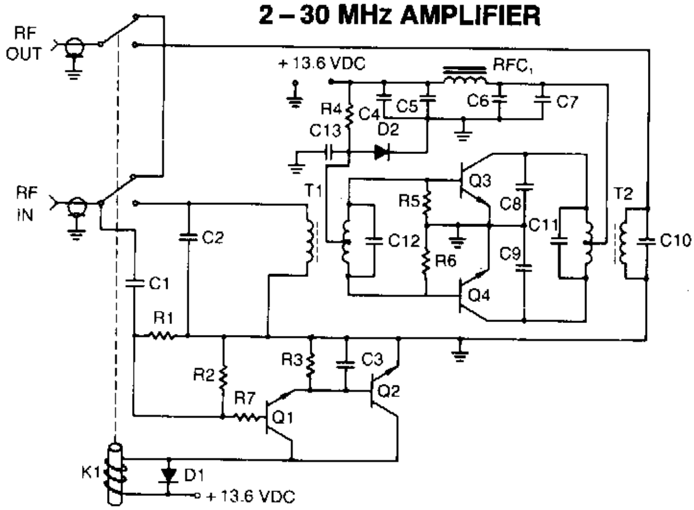


Fig. 86-2

This circuit design is a class A amplifier employing both ac and dc feedback. Bias is stabilized at 15 mA of the collector current using dc feedback from the collector. The ac feedback, from collector to base, and in each of the partially bypassed emitter circuits, compensates for the increase in device gain with decreasing frequency, yielding a flat response over a maximum bandwidth. The amplifier shows a nominal 10-dB power gain from 3 MHz to 1.4 GHz. With only a minimum matching network used at the amplifier input, the input VSWR remains less than 2.5:1 to approximately 1 GHz, while the output VSWR stays under 2:1. Note that a slight degradation in gain flatness and output VSWR occurs with the addition of C6. A more elaborate network design would probably optimize impedance matching, while maintaining gain flatness.

2–30 MHz AMPLIFIER

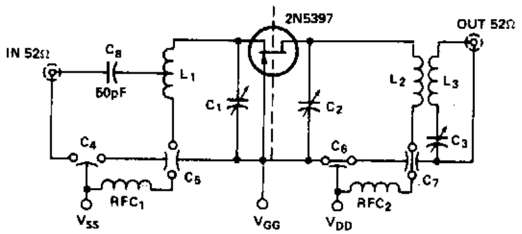


C1—33 pF dipped mica	R—100 KΩ, 0.25 W
C2—18 pF dipped mica	R2—10 KΩ, 0.25 W
C3—10 μF, 35 VDC for AM operation	R3—10 KΩ, 0.25 W
100 μF, 35 VDC for SSB operation	R4—33 Ω, 5 W wirewound
C4—0.1 μF Erie	R5, R6—10 Ω, 0.5 W
C5—10 μF, 35 VDC electrolytic	R7—100 Ω, 0.25 W
C6—1 μF tantalum	RFC1—9 ferroxcube beads on No. 18 AWG wire
C7—0.001 μF Erie disK	D1—1N4001
C8, C9—330 pF dipped mica	D2—1N4997
C10—24 pF dipped mica	Q1, Q2—2N4401
C11—910 pF dipped mica	Q3, Q4—MRF454
C12—1100 pF dipped mica	T1, T2—16:1 transformers
C13—500 μF, 3 VDC electrolytic	K1—Potter & Brumfield KT11A 12 VDC relay or equivalent

MICROWAVES AND RF ***Fig. 86-3***

This amplifier provides 140-W PEP nominal output power when supplied with input levels as low as 3 W. Both input and output transformers have a 4:1 turn ratio and a 16:1 impedance ratio to achieve low input VSWR across the band with high-saturation capability.

450-MHz COMMON-GATE AMPLIFIER



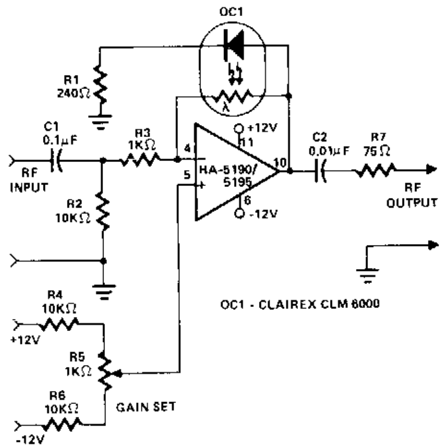
V_{DG} = 10V
I_D = 10mA
NF Typ 3.2dB
G_{ps} Typ 10dB

C_1, C_2, C_3 – 0.8 – 12pF Johanson type 2950
C_4, C_5, C_6, C_7 – 1000pF Allen-Bradley type SS5D
$RFC_{1,2}$ – 0.15μH Delevan type 1537-00
L_1 – 1.5" long; #16 copper
L_2 – 1.2" long; #16 copper
L_3 – 2.0" long; #22 copper enamel, loosely coupled to L_2, 0.75" spacing

This is a low noise, 3-dB typical NF, amplifier with about 10-dB gain at 450–470 MHz for VHF two-way applications.

 Fig. 86-4

RF WIDEBAND ADJUSTABLE AGC AMPLIFIER



HARRIS

Fig. 86-5

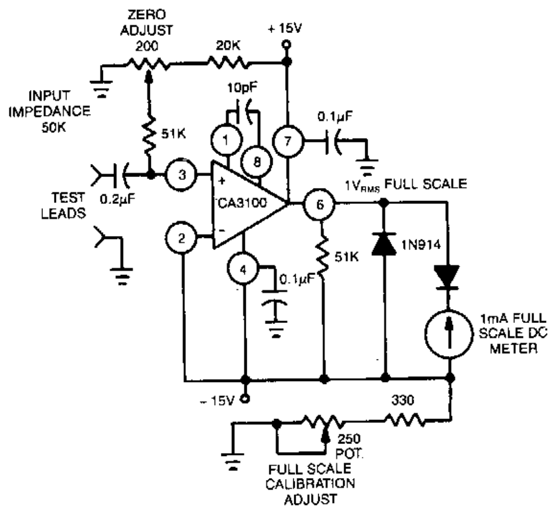
This circuit functions as a wideband adjustable AGC amplifier. With an effective bandwidth of approximately 10 MHz, it is capable of handling rf input signal frequencies from 3.2 to 10 MHz at levels ranging from 40 mV up to 3 V pk-pk.

AGC action is achieved by using optocoupler/isolater OC1 as part of the gain control-feedback loop. In operation, the positive peaks of the amplified output signal drive the OC1 LED into a conducting state. Since the resistance of the OC1 photosensitive element is inversely proportional to light intensity, the higher the signal level, the lower the feedback resistance to the op amp inverting input. The greater negative feedback lowers stage gain. Any changes in gain occur smoothly because the inherent memory characteristic of the photoresistor acts to integrate the peak signal inputs. In practice, the stage gain is adjusted automatically to where the output signal positive peaks are approximately one diode drop above ground.

Gain set control R5 applies a fixed dc bias to the op amp noninverting input, thus establishing the steady state-zero input signal current through the OC1 LED and determining the signal level at which AGC action begins.

The effective AGC range depends on a number of factors, including individual device characteristics, the nature of the rf drive signal, the initial setting for R5, et al. Theoretically, the AGC range can be as high as 4000:1 for a perfect op amp because the OC1 photoresistor can vary in value from 1 MΩ with the LED dark to 250 Ω with the LED fully on.

1-MHz METER-DRIVER AMPLIFIER



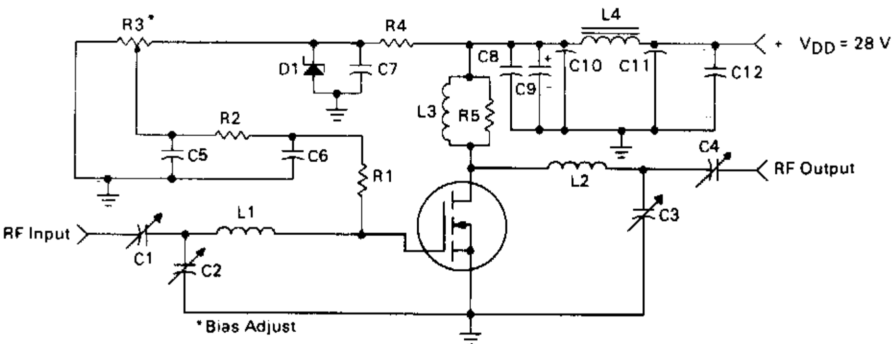
ALL RESISTANCE VALUES ARE IN OHMS

GE/RCA

Fig. 86-6

This circuit uses the CA3100 BiMOS op amp to drive a 1-mA meter movement to full scale with 1-V rms input.

5-W 150-MHz AMPLIFIER



C1, C4 — Arco 406, 15-115 pF
C2 — Arco 403, 3-35 pF
C3 — Arco 402, 1.5-20 pF
C5, C6, C7, C8, C12 — 0.1 μF Erie Redcap
C9 — 10 μF, 50 V
C10, C11 — 680 pF Feedthru
D1 — 1N5925A Motorola Zener
L1 — 3 Turns, 0.310" ID, #18 AWG Enamel, 0.2" Long
L2 — 3-1/2 Turns, 0.310" ID, #18 AWG Enamel, 0.25" Long
L3 — 20 Turns, #20 AWG Enamel Wound on R5
L4 — Ferroxcube VK-200 — 19/4B
R1 — 68 Ω, 1.0 W Thin Film
R2 — 10 kΩ, 1/4 W
R3 — 10 Turns, 10 kΩ Beckman Instruments 8108
R4 — 1.8 kΩ, 1/2 W
R5 — 1.0 MΩ, 2.0 W Carbon
Board — G10, 62 mils

Fig. 86-7

This circuit utilizes the MRF123 TMOS power FET. The MRF134 is a very high gain FET that is potentially unstable at both VHF and UHF frequencies. Note that a 68-Ω input loading resistor has been utilized to enhance stability. This amplifier has a gain of 14 dB and a drain efficiency of 55%.

UHF-TV PREAMPLIFIER

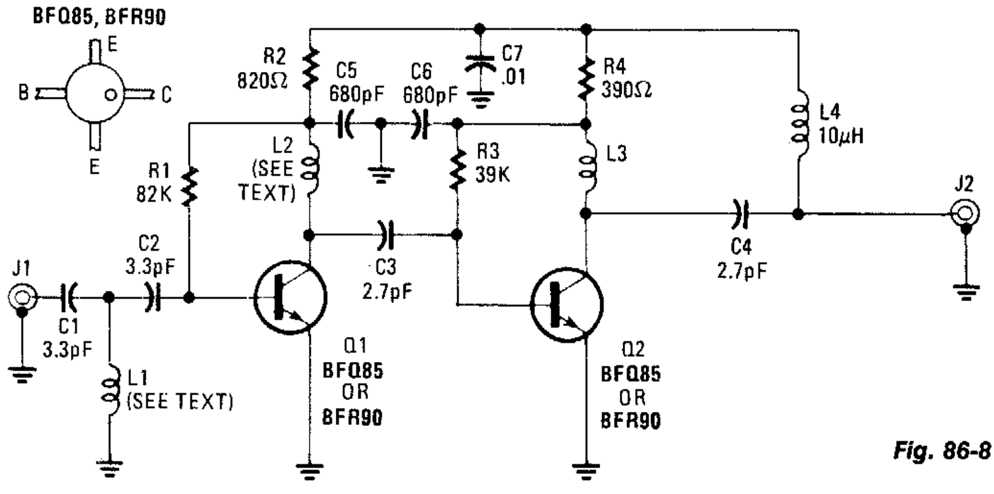
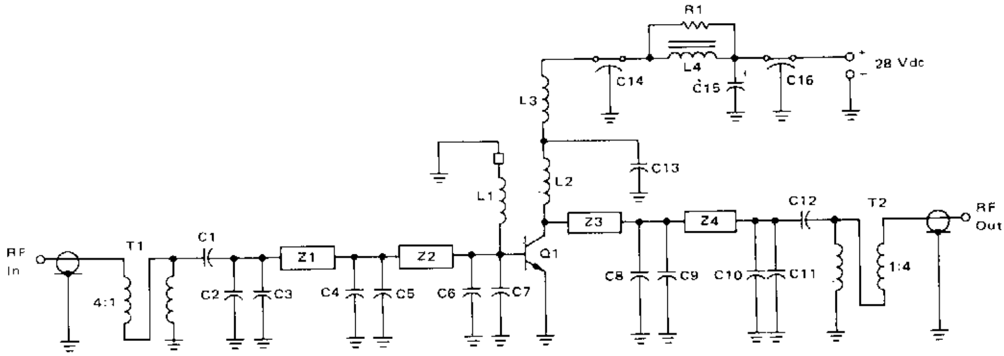


Fig. 86-8

An inexpensive, antenna-mounted, UHF-TV preamplifier can add more than 25 dB of gain. The first stage of the preamp is biased for optimum noise, the second stage for optimum gain. L1, L2 strip line ≈ λ/8 part of PC board.

60-W 225 – 400 MHz AMPLIFIER



C1 – 63 pF Chip
C2, C8 – 27 pF Chip
C3 – 24 pF Chip
C4 – 15 pF Chip
C5, C9 – 30 pF Chip
C6, C7 – 50 pF Chip
C10 10 pF Chip
C11 – 5.1 pF Chip
C12 – 150 pF Chip
C13 – 270 pF Chip
C14, C16 680 pF Feedthru
C15 – 1.0 μF 50 V Tantalum

All Chip Capacitors are 100 mil TDK-ACI Co, Style FC282 BAG
L1 – 0.15 μH Molded Choke with Ferroxcube Bead #56-590-65/4B on ground end of coil
L2 – 1 Turn #22 AWG, 1/8" ID
L3 – 0.15 μH Molded Choke
L4 – Ferroxcube VK-200-19/4B
Q1 – 2N6439
R1 – 10 Ω 2 Watt
T1, T2 – 25 Ω Subminiature Coax (Type UT25) 2.25 inches (57.15 mm) long

Z1 – Microstrip Line 800 mils L X 225 mils W 20.32 mm L X 5.715 mm W
Z2 – Microstrip Line 200 mils L X 225 mils W 5.08 mm L X 5.715 mm W
Z3, Z4 – Microstrip Line 550 mils L X 125 mils W 13.97 mm L X 3.175 mm W
Board – 0.031" (0.787 mm) Glass Teflon $\epsilon_r = 2.56$

2N6439 60 Watt Building Block 225 – 400 MHz

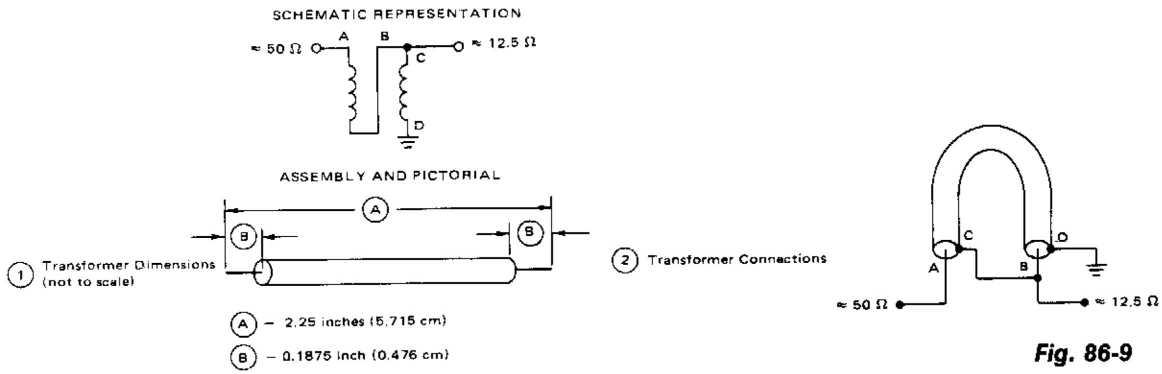


Fig. 86-9

Construction Details of the 4:1 Unbalanced to Unbalanced Transformers

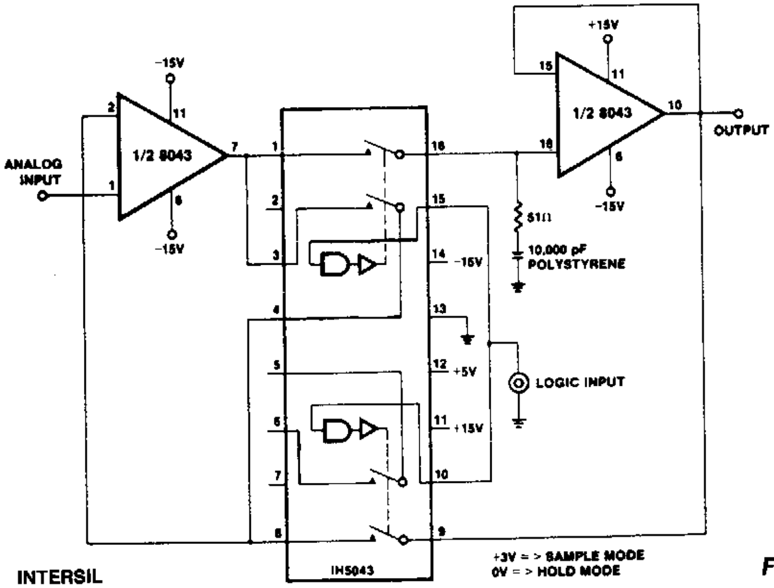
This 60-W, 28-V broadband amplifier covers the 225 – 400 MHz military communications band. The amplifier may be used singly as a 60-W output stage in a 225 – 400 MHz transmitter, or by using two of these amplifiers combined with quadrature couplers, a 100-W output amplifier stage can be constructed. The circuit is designed to be driven from a 50-Ω source and work into a nominal 50-Ω load. The input network consists of two microstrip L-sections composed of Z1, Z2, and C2 through C6. C1 serves as a dc-blocking capacitor. A 4:1 impedance ratio coaxial transformer T1 completes the input matching network. L1 and a ferrite bead serve as a base decoupling choke. The output circuit consists of shunt inductor L2 at the collector, followed by two microstrip L-sections composed of Z3, Z4, and C8 through C11. C12 serves as a dc blocking capacitor, and is followed by another 4:1 impedance ratio coaxial transformer. Collector decoupling is accomplished through the use of L3, L4, C14, C15, C16, and R1.

87

Sample-and-Hold Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SAMPLE-AND-HOLD



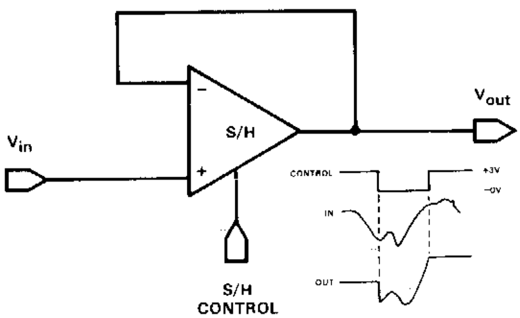
INTERSIL

Fig. 87-1

Two important properties of the 8043 are used to advantage in this circuit. The low input bias currents give rise to slow output decay rates (droop) in the hold mode, while the high slew rate at 6 V/μs improves the tracking speed and the response time of the circuit. The upper waveform is the input 10 V/div, the lower waveform the output 5 V/div. The logic input is high.

The center waveform is the analog input, a ramp moving at about 67 V/ms, the lower waveform is the logic input to the sample-and-hold; a logic 1 initiates the sample mode. The upper waveform is the output, displaced by about one scope division 2 V from the input to avoid superimposing traces. The hold mode, during which the output remains constant, is clearly visible. At the beginning of a sample period, the output takes about 8 μs to catch up with the input, after which it tracks, until the next hold period.

BASIC TRACK-AND-HOLD/SAMPLE-AND-HOLD

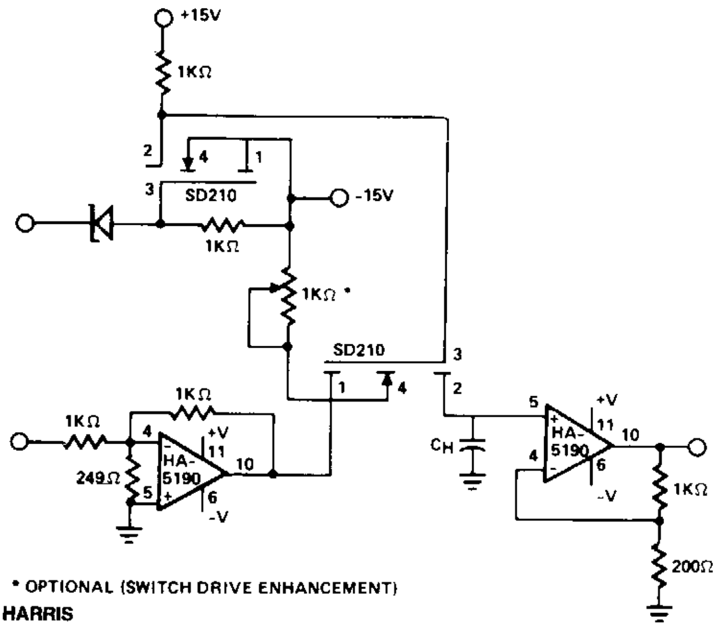


HARRIS

Fig. 87-2

Feedback is the same as a conventional op-amp voltage follower which yields a unity-gain, noninverting output. This hookup also has a very high input impedance. The only difference between a track-and-hold and a sample-and-hold is the time period during which the switch is closed. In track-and-hold operation, the switch is closed for a relatively long period; the output signal might change appreciably and would hold the level present at the instant the switch is opened. In sample-and-hold operation, the switch is closed only for the time necessary to fully charge the holding capacitor.

HIGH-SPEED SAMPLE-AND-HOLD



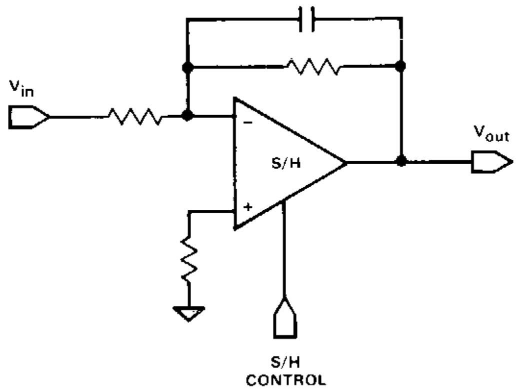
* OPTIONAL (SWITCH DRIVE ENHANCEMENT)

HARRIS

Fig. 87-3

This circuit uses the speed and drive capability of the HA-5190 coupled with two high-speed DMOS FET switches. The input amplifier is allowed to operate at a gain of −5, although the overall circuit gain is unity. Acquisition times of less than 100 ns to 0.1% of a 1-V input step are possible. Drift current can be appreciably reduced by using FET input buffers on the output stage of the sample-and-hold.

FILTERED SAMPLE-AND-HOLD

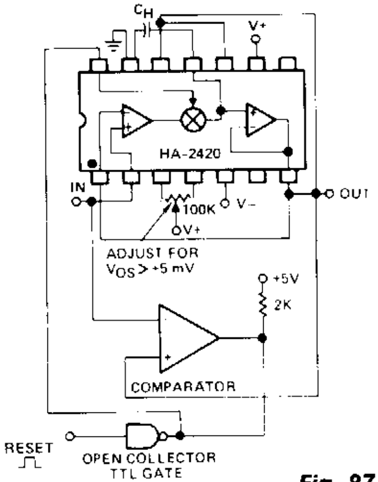


It is often required that a signal be filtered prior to sampling. This can be accomplished with only one device. Use any of the inverting and noninverting filters that can be built with op amps. However, it is necessary that the sampling switch be closed for a sufficient time for the filter to settle when active filter types are connected around the device.

HARRIS

Fig. 87-4

SAMPLE-AND-HOLD

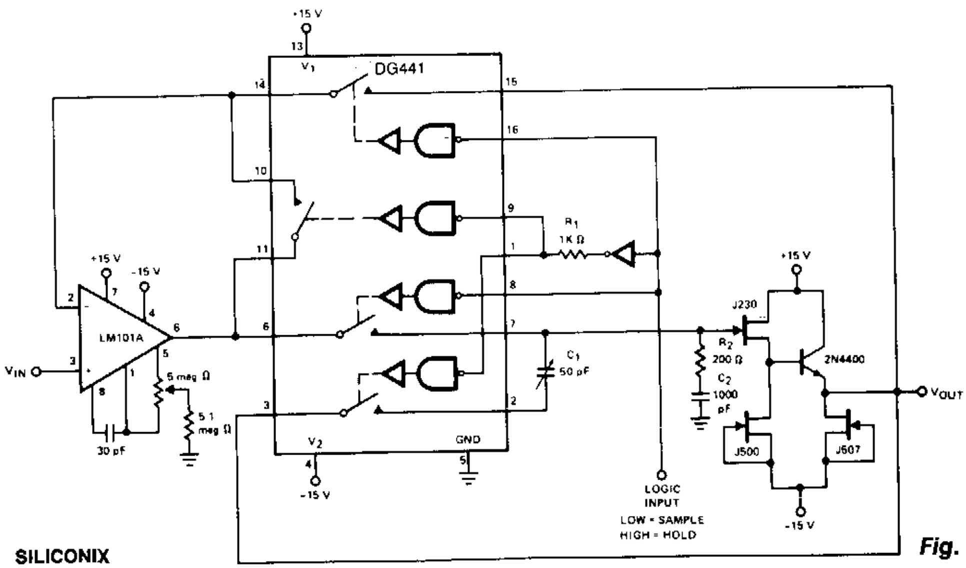


HARRIS

Fig. 87-5

The sample-and-hold function has often been accomplished with separate analog switches and op amps. These designs always involve performance tradeoffs between acquisition time, charge injection, and droop rate. The HA-242-/2425 monolithic sample-and-hold, has many better tradeoffs, and usually a lower total cost than the other approaches. The switching element is a complementary bipolar circuit with feedback, which allows high charging currents of 30 mA, a low charge injection of 10 pC, and an ultra-low off leakage current of 5 pA; a combination that is not approached in any other electronic switch. These factors make it also superior as an integrator reset switch, or as a precision peak detector.

SAMPLE-AND-HOLD

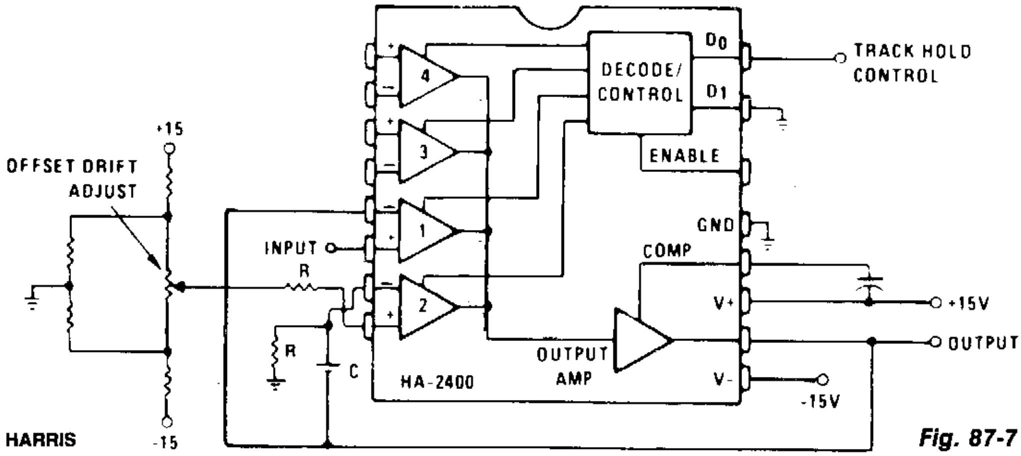


SILICONIX

Fig. 87-6

The LM101A provides gain and buffers the input from storage capacitor C2. R2 adds a zero in the open loop response to compensate for the pole caused by the switch resistance and C2, improving the closed-loop stability. R1 provides a slight delay in the digital drive to pins 1 and 9. C1 provides cancellation of coupled charge, keeping the sample-and-hold offset below 5 mV over the analog signal range of −10 through +10 V. Aperture time is typically 1 μs, the switching time of the DG441. Acquisition time is 25 μs, but this can be improved by using a faster slewing op amp. Droop rate is typically less than 5 mV/s at 25°C.

TRACK-AND-HOLD/SAMPLE-AND-HOLD

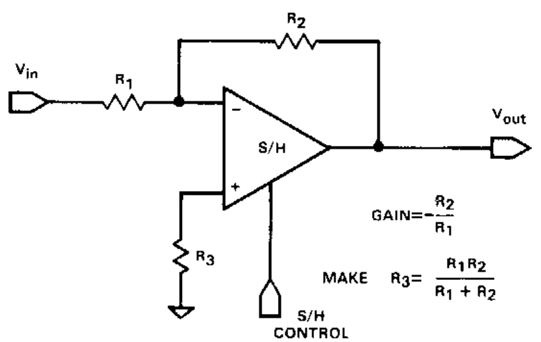


HARRIS

Fig. 87-7

Channel 1 is wired as a voltage follower and is turned on during the track/sample time. If the product of $R \times C$ is sufficiently short compared to the period of maximum output frequency, or sample time; C will charge to the output level. Channel 2 is an integrator with zero input signal. When channel 2 is then turned on, the output will remain at the voltage across C.

INVERTING SAMPLE-AND-HOLD



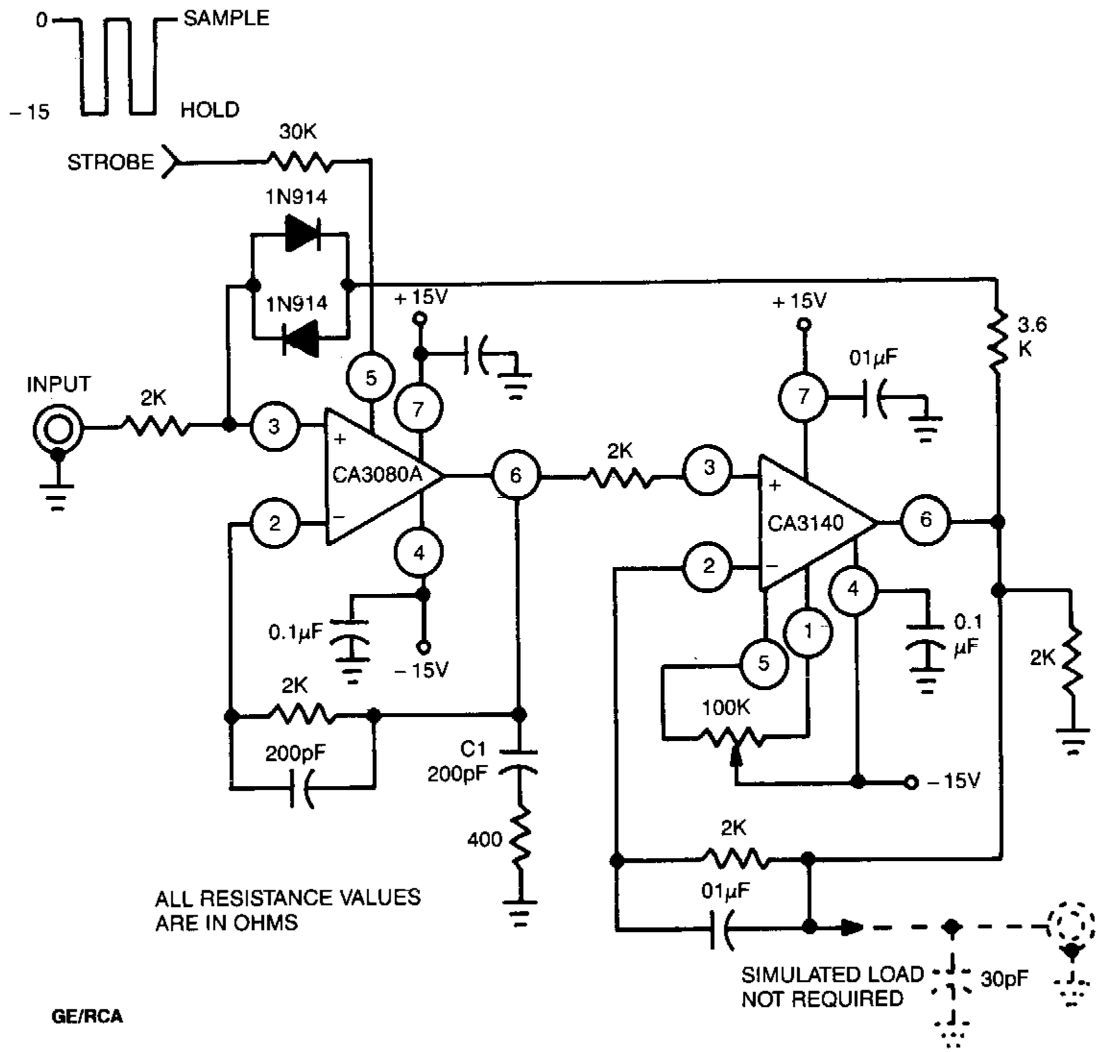
HARRIS

Fig. 87-8

This illustrates another application in which the hookup versatility of a sample-and-hold often eliminates the need for a separate op amp and a sample-and-hold module. This hookup will have a somewhat higher input-to-output feedthrough during hold than the noninverting connection, since the output impedance is an open-loop value during hold. The feedthrough will

$$\frac{V_{IN}\,R_0}{R_1 + R_2 + R_0}$$

SAMPLE-AND-HOLD



GE/RCA

Fig. 87-9

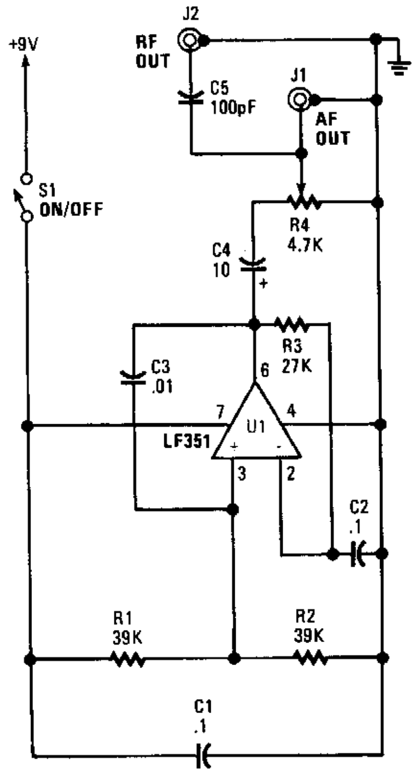
This circuit uses a CA3140 BiMOS op amp as the readout amplifier for the storage capacitor C1, and a CA3080A variable op amp as input buffer amplifier and low feedthrough transmission switch. Offset nulling is accomplished with the CA3140.

88

Signal Injectors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

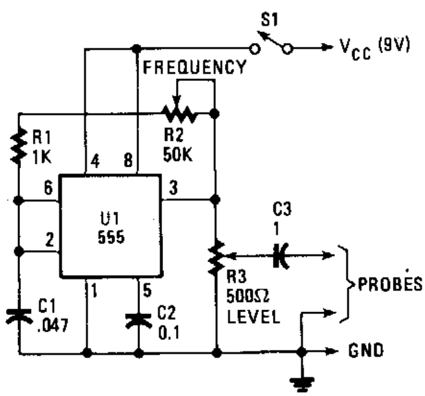
SIGNAL INJECTOR



This unit is a single oscillator built around an LF351 JFET-input op amp. Resistors R1 and R2 bias the noninverting input while R3 biases the inverting input from the output. This layout provides 100% negative feedback, but the decoupling caused by C2 gives reduced feedback and high-voltage gain when dealing with audio frequencies. The fundamental operating frequency is about 800 Hz. Potentiometer R4 is the output-level control. To use it start at the speaker. If no tone is heard, move back to the amplifer input, and listen for the tone. Still if no tone is heard, continue backtracking from the output to the input, covering all stages in between. The stage where the signal is lost is the one that is not operating.

POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 88-1*

SIGNAL INJECTOR



The unit provides a square-wave output that is rich in harmonic content. The circuit's output frequency can be varied from 50 Hz to 15 kHz. The heart of the circuit is a 555 astable connected in its equal mark/space mode. The frequency is controlled by potentiometer R2 and capacitor C1. Resistor R3 controls the output level with the output ac-coupled through C3.

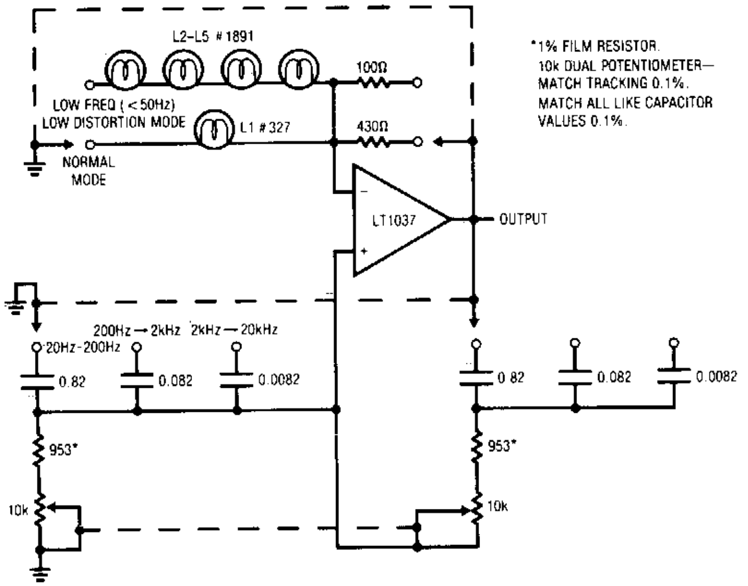
POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 88-2*

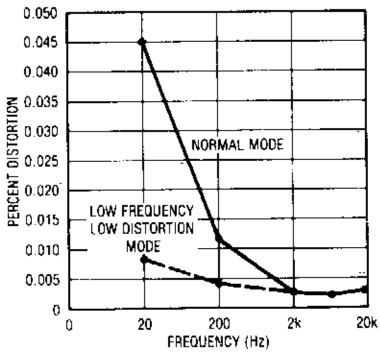
89

Sine-Wave Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

LOW-DISTORTION THERMALLY STABILIZED WIEN-BRIDGE OSCILLATOR





Oscillator Distortion vs Frequency

LINEAR TECHNOLOGY CORP. ***Fig. 89-1***

A variable Wien bridge provides frequency tuning from 20 Hz to 20 kHz. Gain control comes from the positive temperature coefficient of the lamp. When power is applied, the lamp is at a low resistance value, the gain is high, and oscillation amplitude builds. The lamp's gain-regulating behavior is flat within 0.25 dB over the 20 Hz – 20 kHz range of the circuit. Distortion is below 0.003%. At low frequencies, the thermal time constant of the small normal-mode lamp begins to introduce distortion levels about 0.01%. This is because of ***hunting*** when the oscillator's frequency approaches the lamp's thermal time constant. This effect can be eliminated, at the expense of reduced output amplitude and longer amplitude settling time, by switching to the low-frequency, low-distortion mode. The four large lamps give a longer thermal time constant, and distortion is reduced.

SINGLE-SUPPLY WIEN-BRIDGE OSCILLATOR

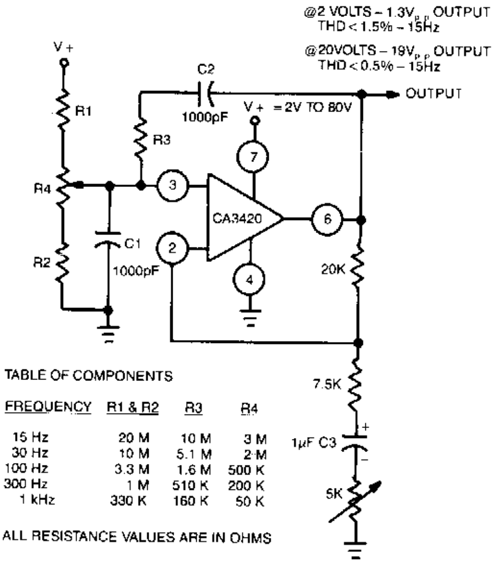


TABLE OF COMPONENTS

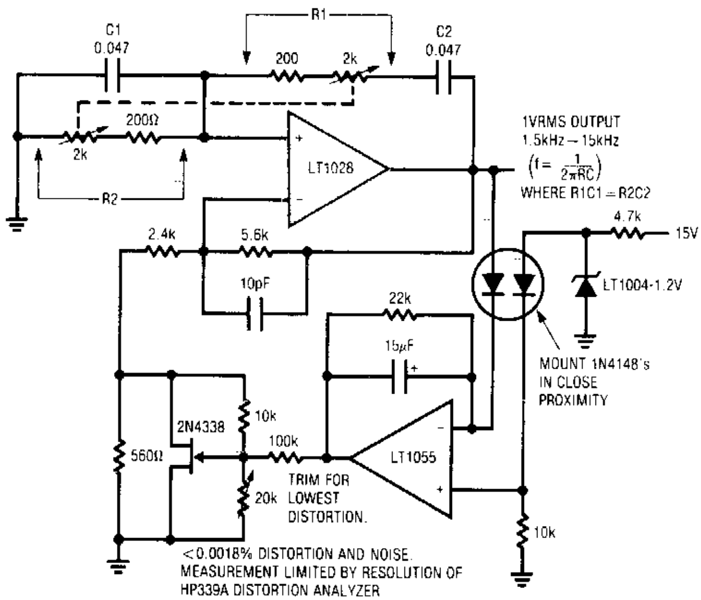
FREQUENCY	R1 & R2	R3	R4
15 Hz	20 M	10 M	3 M
30 Hz	10 M	5.1 M	2 M
100 Hz	3.3 M	1.6 M	500 K
300 Hz	1 M	510 K	200 K
1 kHz	330 K	160 K	50 K

ALL RESISTANCE VALUES ARE IN OHMS

GE/RCA ***Fig. 89-2***

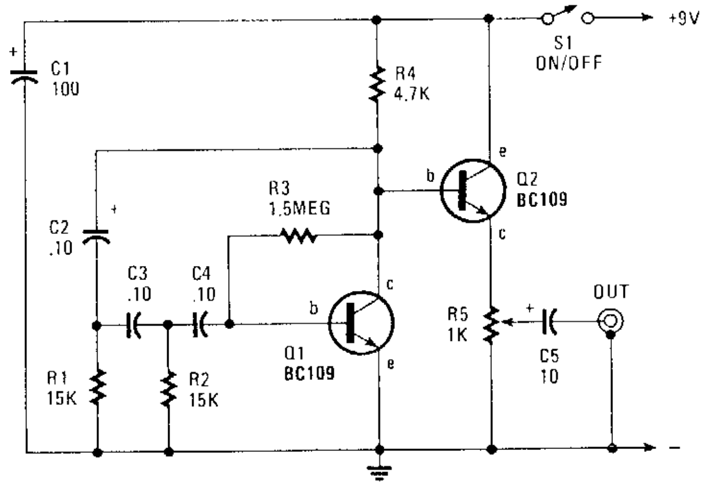
The adjustment of R4 contributes to the comparatively symmetrical output transfer characteristic of the CA3420 BiMOS op amp. To extend the lower operating frequency, remove C3 and use a dual supply.

SUPER-LOW-DISTORTION VARIABLE SINE-WAVE OSCILLATOR



LINEAR TECHNOLOGY CORP. ***Fig. 89-3***

AUDIO GENERATOR



HANDS-ON ELECTRONICS

Fig. 89-4

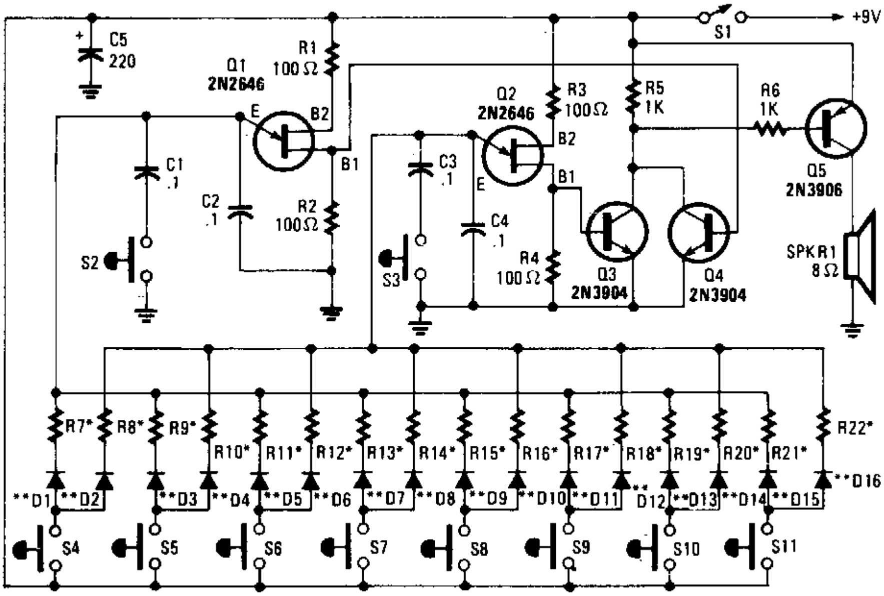
This circuit produces a sinusoidal output of about 8 V pk-pk, which can be varied down to zero, at about 500 Hz. The signal is generated by a phase-shift oscillator.

90

Sirens, Warblers, and Wailers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

ELECTRONIC BAGPIPE



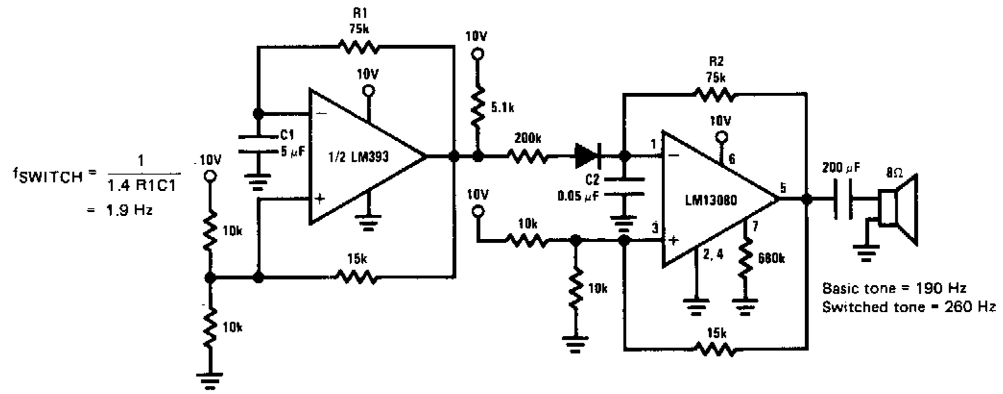
* SEE TEXT
**D1–D16 ARE 1N914

POPULAR ELECTRONICS *Fig. 90-1*

This circuit mimics the dual-tone drone sound that's produced by the unusual wind instrument. Unijunction transistors Q1 and Q2 are connected in similar audio-oscillator circuits. Each of the oscillator frequencies is determined by one of the two resistors selected by one of the pushbutton switches, S4 through S11. The odd-numbered resistors in R7 to R21, determine the frequency for the Q1 oscillator circuit and the even-numbered resistors in R8 through R22, determine the frequency for Q2's circuit.

When S4 is pressed, the positive supply is connected to both R7 and R8 through isolation diodes D1 and D2, causing both oscillators to operate. A narrow, fast-rising positive pulse is developed at B1 of both Q1 and Q2 for each cycle of operation. Transistors Q3 and Q4 serve as a simple audio mixer, which is used to combine the pulses from each oscillator. The mixed signal at the collectors of Q3 and Q4 is coupled through R6 to the base of Q5, which amplifies and drives an 8-Ω speaker, SPKR1. Switches, S2 and S3 are used to reduce the oscillator's frequency by about 50% when closed, to produce a new group of tones.

TWO-TONE SIREN

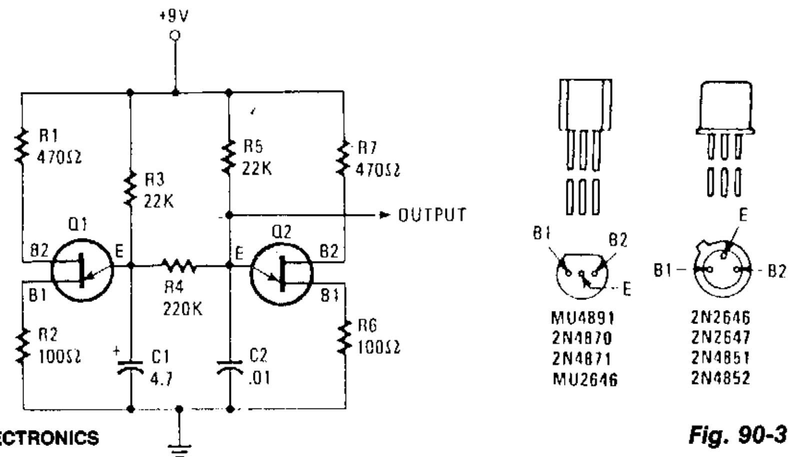


NATIONAL SEMICONDUCTOR

Fig. 90-2

This siren provides a constant audio output, but alternates between two separate tones. The LM13080 is set to oscillate at one basic frequency; this frequency is changed by adding a 200-KΩ charging resistor in parallel with the feedback resistor, R2.

YELPING SIREN

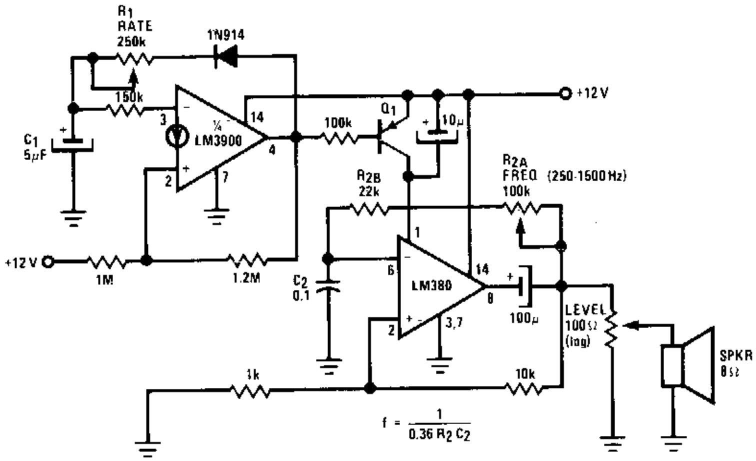


HANDS-ON ELECTRONICS

Fig. 90-3

Unijunction transistors Q1 and Q2 are both connected as relaxation-type, sawtooth oscillators. Transistor Q1 is the low-frequency control oscillator and Q2 is the tone generator. Sawtooth waveforms are produced at the emitter terminals. Without R4 connecting the two emitters, each oscillator operates independently, with its frequency determined mainly by the rc time constant. With the values shown, Q1 operates from 1. to 1.5 Hz and Q2 operates from 400 to 500 Hz. When R4 is connected between the two emitters, it couples the low-frequency sawtooth from Q1 directly across capacitor C2. That coupling causes the frequency of the tone generator to increase, along with the rise in sawtooth voltage from Q1. The tone generator's frequency drops to its lower design value when C1 discharges and produces the falling edge of the sawtooth.

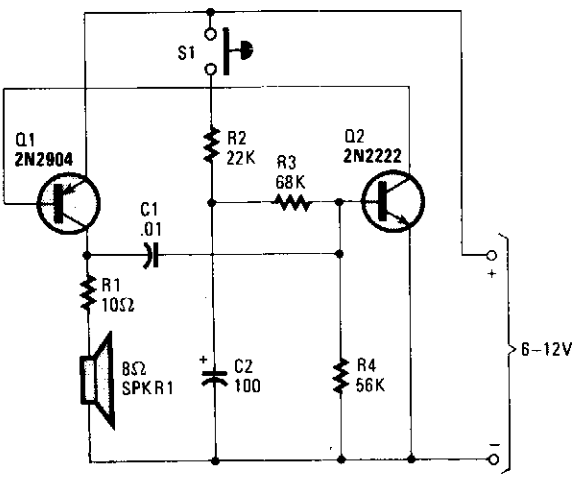
PROGRAMMABLE-FREQUENCY ADJUSTABLE-RATE SIREN



NATIONAL SEMICONDUCTOR CORP. *Fig. 90-4*

The LM380 operates as an astable oscillator with the frequency determined by R2/C2. Adding Q1 and driving its base, with the output of an LM3900 wired as a second astable oscillator, acts to gate the output of the LM380 on and off, at a rate fixed by R1/C1.

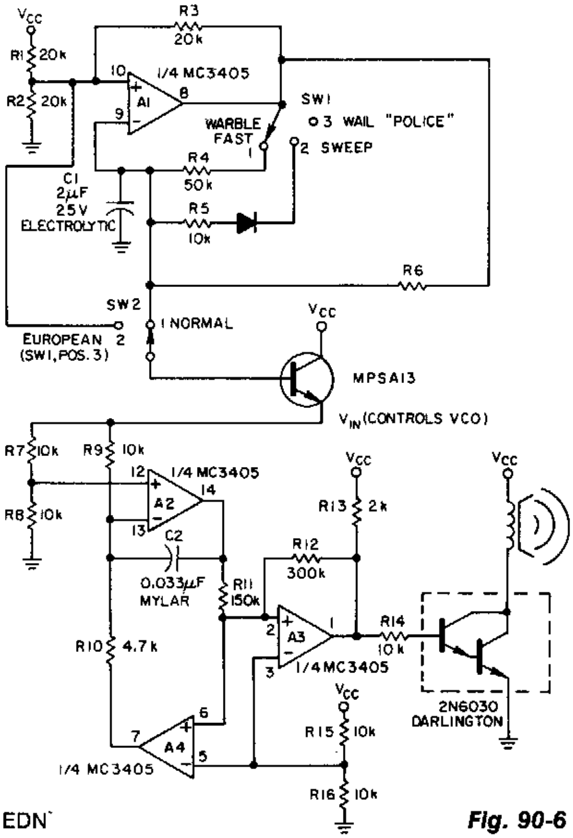
THE WAILING SIREN



HANDS-ON ELECTRONICS *Fig. 90-5*

Transistors Q1 and Q2, with feedback provided via C1 from the collector of Q1 to the base of Q2, forms a voltage-controlled oscillator (VCO). Depending on the voltage applied to Q2's base, the VCO frequency ranges from around 60 Hz to 7.5 kHz. The instantaneous voltage applied to the base of Q2 is determined by the values of C2, R2, R3, and R4. When pushbutton switch S1 is closed, C2 charges fairly rapidly to the maximum supply voltage through R2, a 22-KΩ fixed resistor. That causes the siren sound to rise rapidly to its highest frequency. When the button is released, the capacitor discharges through R3 and R4 with a combined resistance of 124 KΩ, causing the siren sound to decay from a high-pitched wail to a low growl. If you want to experiment with the pitch of the sound at its highest frequency, try different values for C1. Increase its value for lower notes, and decrease it for higher ones. Different values for R2 will change the attack time. A 100-KΩ resistor provides equal attack and decay times. The way you handle the pushbutton varies the effect.

LINEAR IC SIREN

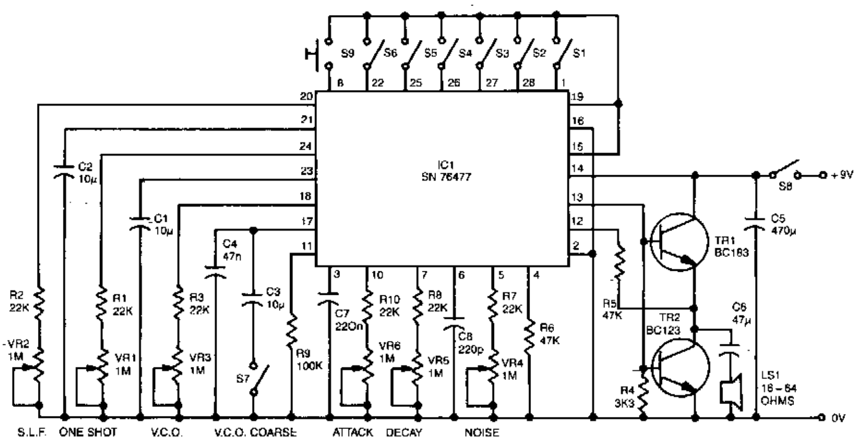


EDN

Fig. 90-6

A low-frequency, op-amp oscillator and a VCO, both configured from a single MC3405 dual op amp and dual comparator, are the major components in a siren circuit that can be made to produce various warbles and wails, or serve as an audio sweep generator. The only other active components needed are an MPS A13 small-signal transistor and a 2N6030 power Darlington transistor.

SUPER SOUND GENERATOR

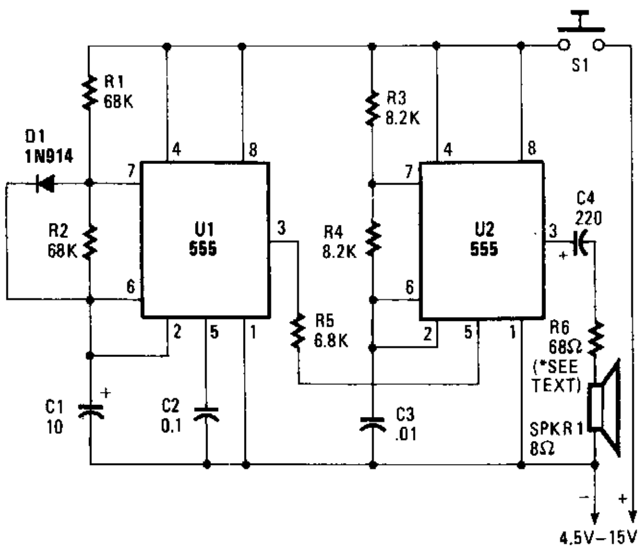


EVERDAY ELECTRONICS

Fig. 90-7

Six preset controls and seven selector switches enable a vast range of different sounds to be produced and altered at will. Such sounds as steam trains chuffing, helicopters flying, bird chirping, and machine guns firing are possible, as well as the usual police sirens, phaser guns, and bomb explosions. The circuit incorporates an amplifier giving 150-mW output into a small loudspeaker. Alternatively, a separate amplifier system can be used for disco effects, car alarms, etc. Continuous or one-shot sounds are possible. For one-shot sounds, a push-button switch is provided, which can also be used to turn continuous sounds on and off. A single IC, SN76477, provides all of the sound generation circuits.

HEE-HAW SIREN



HANDS-ON ELECTRONICS

Fig. 90-8

A pair of timer IC's are the heart of a circuit that simulates the warbling hee-haw of a British police siren. One of the 555 timers, U2, is wired as an astable multivibrator operating at about 900 Hz. The other, U1, operates at approximately 1 Hz. Its output at pin 3 is a square wave with a 50% duty cycle—on and off cycles of about 0.5 second each. The output of U1 is applied to pin 5, the control-voltage terminal of U2. The frequency of the 555 timer IC is relatively independent of supply voltage, but can be varied over a fairly wide range by applying a variable voltage between pin 5 and ground. When U1's output becomes high, U2 operates at about 800 kHz. That switching between two frequencies produces the warbling hee-haw signal.

ELECTRONIC SIREN

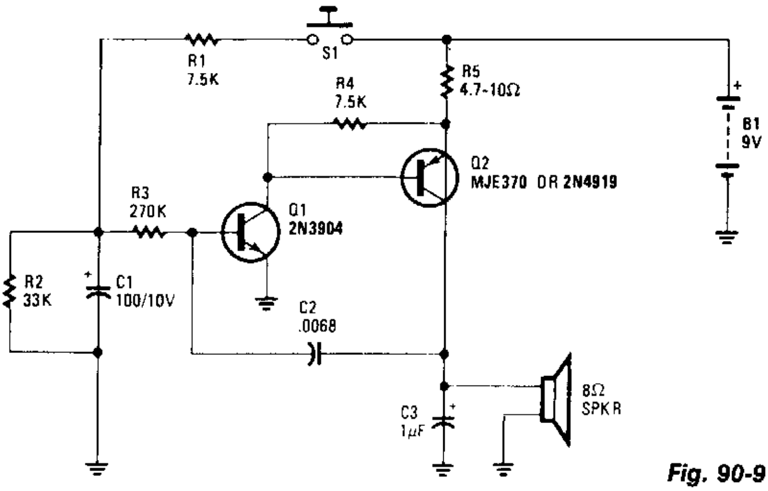
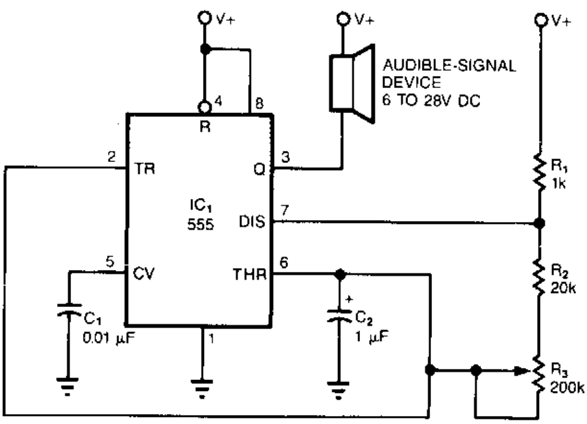


Fig. 90-9

The wailing sound of a siren is generated by a VFO consisting of Q1 and Q2. Capacitor C2 provides the feedback for the oscillator. The frequency of the oscillator is varied by the voltage applied to the base of Q1 through R3. When switch S1 is closed, capacitor C1 charges, thus increasing the oscillator frequency. When S1 is released, capacitor C1 discharges, and the oscillator frequency decreases. Capacitor C3 limits the maximum oscillator frequency. The average battery current drain is about 15 mA.

555 BEEP TRANSFORMER

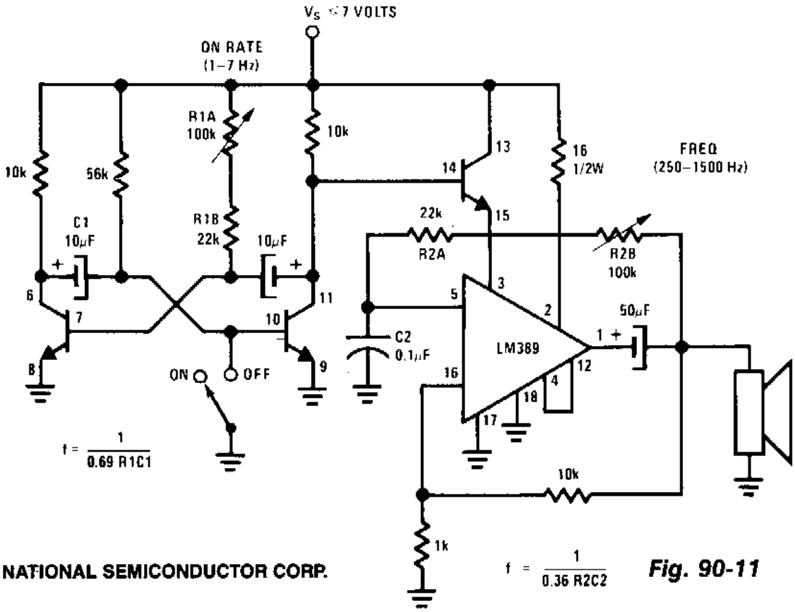


EDN

Fig. 90-10

The simple circuit transforms the steady beep of an audible-signal device, such as a Mallory sonalert, into a distinctive warble or chirp. The value of C2 determines just what tone color you'll get. With the 1-μF value shown, the circuit produces a warble similar to the ring tone of an inexpensive phone. A 10-μF value produces a chirp similar to a truck's back-up alarm. One elaboration of this circuit would be to use the second section of a 555 timer to drive a piezoelectric transducer instead of a sonalert; that modification would vary the tone's pitch, as well as the chirp rate.

SIREN



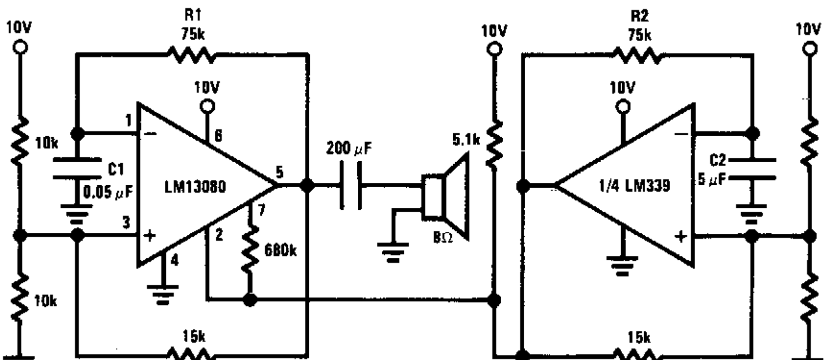
NATIONAL SEMICONDUCTOR CORP.

Fig. 90-11

This circuit uses one of the LM389 transistors to gate the power amplifier on and off by applying the muting technique. The other transistors form a cross-coupled multivibrator circuit that controls the rate of the square-wave oscillator. The power amplifier is used as the square-wave oscillator with individual frequency adjust provided by potentiometer R2B.

TWO-STATE SIREN

$$f_{AUDIO} = \frac{1}{1.4\,R1C1} = 190\text{ Hz}$$



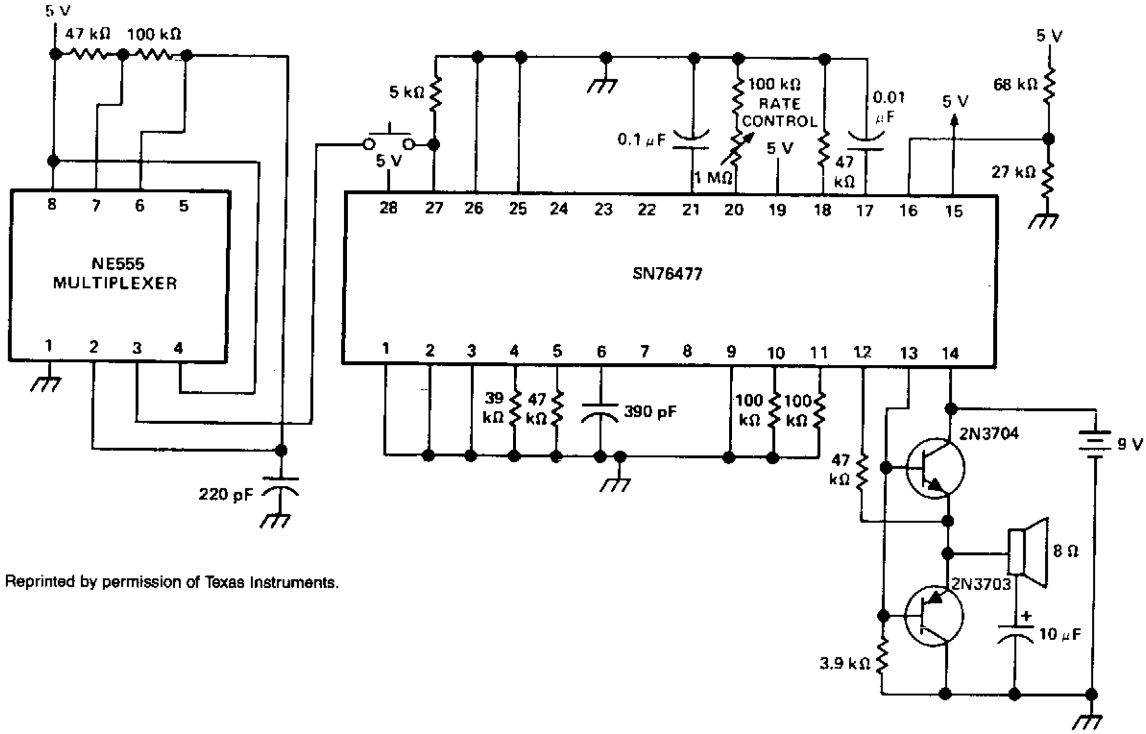
$$f_{SWITCH} = \frac{1}{1.4\,R2C2} = 1.9\text{ Hz}$$

NATIONAL SEMICONDUCTOR CORP.

Fig. 90-12

This is a two-state or on/off-type siren where the LM13080 oscillates at an audio frequency and drives an 8-Ω speaker. The LM339 acts as a switch which controls the audio burst rate.

STEAM TRAIN WITH WHISTLE



Reprinted by permission of Texas Instruments.

Fig. 90-13

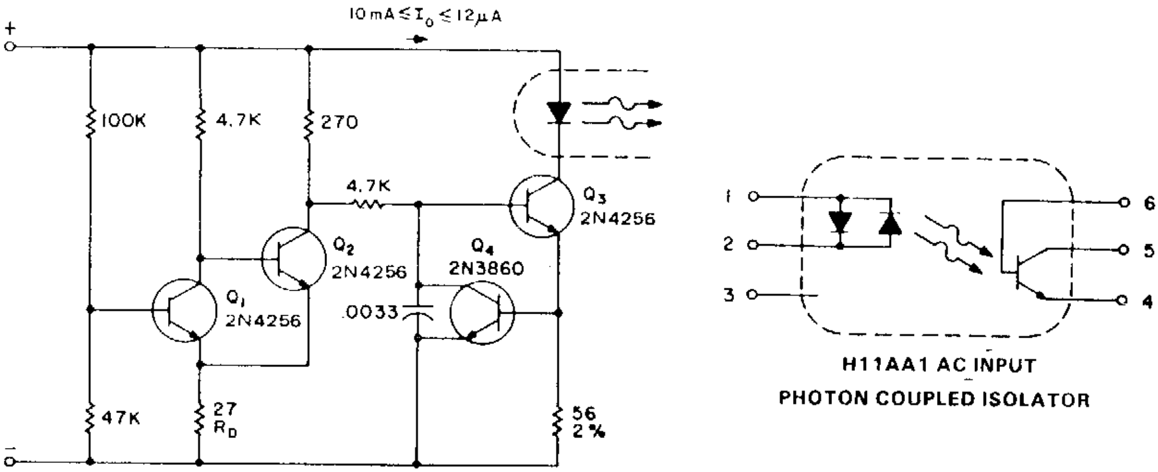
91

Solid-State Relay Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Ac Solid-State Relays

AC SOLID-STATE RELAYS



GE

Fig. 91-1

In the case where analog signals are being used as the logic control, hysteresis from a Schmitt-trigger input can be used to prevent half-wave power output. The circuit operation is as follows: at low input voltages, Q1 is biased in the off state. Q2 conducts and biases Q3, and the IRED turns off. When the base of Q1 reaches the biasing voltage of 0.6 V, plus the drop across R_D, Q1 turns on. Q3 is then supplied base drive, and the solid-state relay input will be activated. The combination of Q3 and Q4 acts as a constant-current source to the IRED. In order to turn-off Q3, the base drive must be reduced to pull it out of saturation. Because Q2 is in the off-state as the signal is reduced, Q1 will now stay on to a base bias-voltage lowered by the change in the drop across R_D. With these values, the highest turn-off voltage is 1.0 V, while turn-on will be at less than the 4.1 V supplied to the circuit.

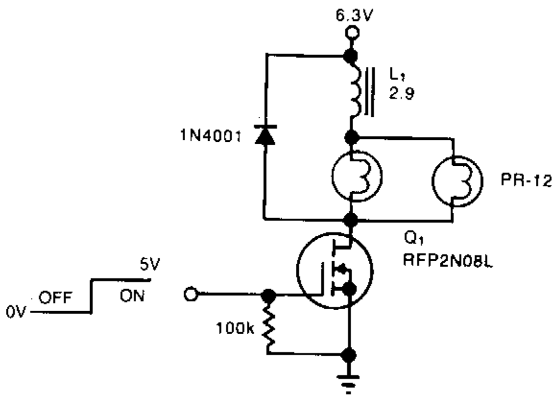
For ac or bipolar input signals, there are several possible connections. If only positive signals are set to activate the relay, a diode, such as the A14, can be connected in parallel to protect the IRED from reverse voltage damage, since its specified peak reverse voltage capability is approximately 3 V. If ac signals are being used, or if activation is to be polarity insensitive, a H11AA coupler, which contains two LEDs in antiparallel connection, can be used. For high-input voltage designs, or for any easy means of converting a dc input relay to ac, a full-wave diode bridge can be used to bias the IRED.

92

Solenoid Drivers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

POWER-CONSUMPTION LIMITER



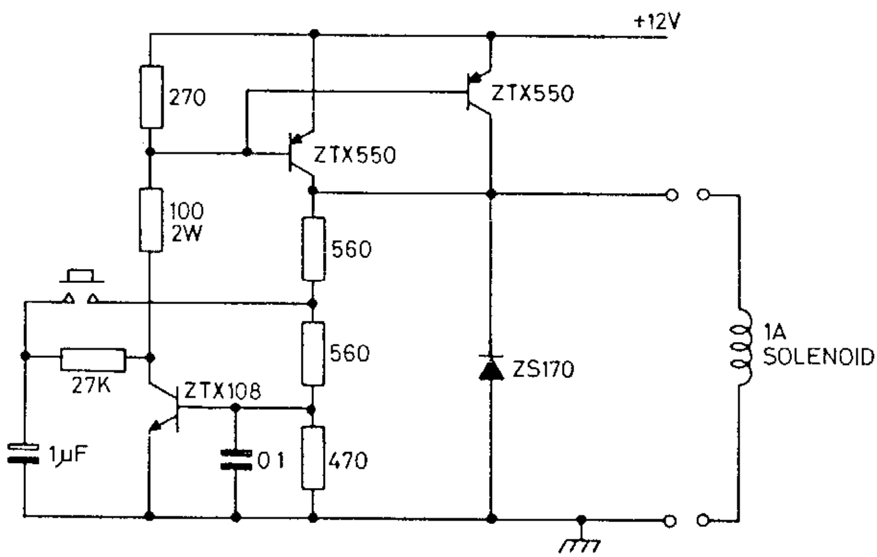
EDN

Fig. 92-1

A simple solenoid driver uses incandescent lamp filaments as on-indicators to limit power consumption. High magnetic reluctance (opposition to flux) in the coil of an armature-driven device, such as a solenoid or relay, calls for a surge of activation current, followed by a lower dc level to remain on, since surge to on-current ratio is typically 5:1. The cold filament allows a surge of coil-activation current to pass through; as the filament heats up, it throttles the current to a more reasonable hold value. The solenoid driver circuit offers these features:

- 5-V logic swings turn the power-MOSFET switch, Q1, fully on and off.
- Two low-cost flashlight lamps, in parallel, handle the peak current. Because their dc current is only 50% of peak and because they operate at 60% of their rated voltage, the lamps have an operating life of 12,000 hours. Further, the lamp filaments' positive temperature coefficients raise each filament's resistance. This rise in resistance eliminates current-hogging problems and provides short-circuit protection.
- The steady-state on-current is 700 mA, vs. 1700 mA without the lamps.
- A 4.6-V min supply rating allows battery operation.

12-V LATCH



ZeTeX, formerly FERRANTI

Fig. 92-2

This circuit controls a solenoid by the operation of a single push-button switch. The circuit will supply loads of over 1 A and can be operated up to a maximum speed of once every 0.6 second. When power is first applied to the circuit, the solenoid will always start in its off position. Other features of the circuit are its automatic turn-off, if the load is shorted, and its virtually zero-power consumption when off.

HOLD-CURRENT LIMITER

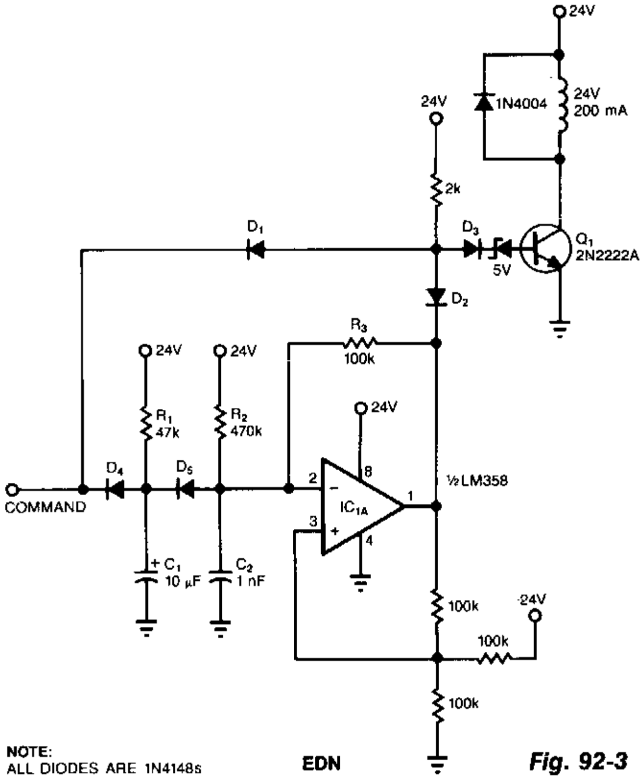


Fig. **92-3**

In many applications, a solenoid driver must first briefly supply a large amount of pull-in current, which quickly actuates the solenoid. Thereafter, the driver must supply a much lower holding current to avoid burning the solenoid out. To avoid using the customary, cumbersome, large capacitors or power-wasting resistors, you can use the switch technique.

As long as the input to the circuit is low, diode D1 holds Q1 off; a low input also prevents the op-amp circuit from oscillating. When the input reaches 24 V, Q1 switches on and pulls in the solenoid. Concurrently, D4 is back-biased, and C1 begins charging up. When C1 charges up, the op-amp circuit begins to oscillate, switching Q1 on and off.

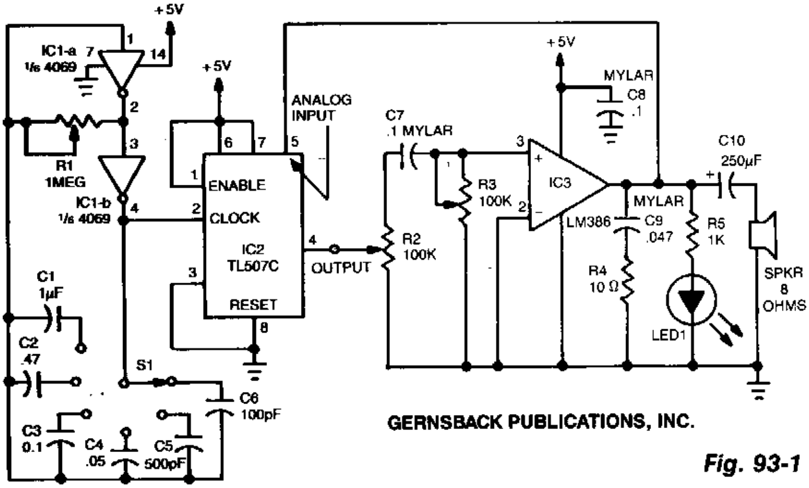
The time constant defined by R1 and C1 determines the length of the period during which the solenoid receives full power. R3 and C2 set the oscillator's frequency, and R2 sets the oscillator's duty cycle. The hold current is directly proportional to the duty cycle. For the components shown, the full-power period is 300 ms, the oscillator's frequency is 3 kHz, and its duty cycle is 50%.

93

Sound Effects

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SOUND-EFFECTS GENERATOR



GERNSBACK PUBLICATIONS, INC.

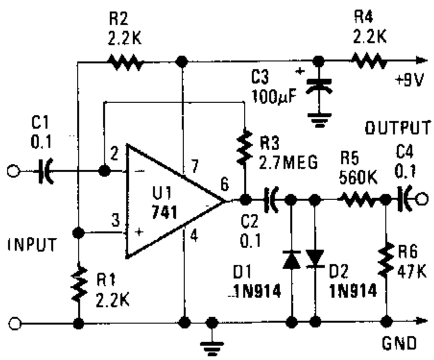
Fig. 93-1

A variable clock-pulse generator is made up of two sections of IC1 (a 4069 CMOS hex inverter), R1, S1, and capacitors C1 through C6. By adjusting R1 and switching one of the capacitors into the circuit, the clock's pulse rate can be varied over a wide range.

The TL507C converts analog signals—in this case the output of IC3, an LM386 audio amplifier—into digital signals. The conversion is accomplished using the single-slope method; it involves comparing an internally generated ramp signal to the analog input signal and a 200-mV reference voltage.

The square-wave output from the a/d converter is fed to IC3 through a network consisting of R2, R3, and C7. Resistor R2 controls the amplitude of the pulses. Resistor R3 and capacitor C7 form a variable tone-control filter and a differentiator circuit that converts a square wave into a spiked waveform. That waveform is amplified by IC3, and the resulting output is fed back into the analog input of IC2, as well as to an 8-Ω speaker. By adjusting R1 and selecting one of the six capacitors with S1—thus varying the clock frequency—and by varying R2 and R3, you can produce many sounds.

FUZZ BOX



HANDS-ON ELECTRONICS

Fig. 93-2

The 741's maximum gain of 20,000 is pushed to nearly 3 million dB, and therefore distorts the output. That distortion provides the fuzz sound. The level is dropped by clipping the two diodes.

CHUG-CHUG

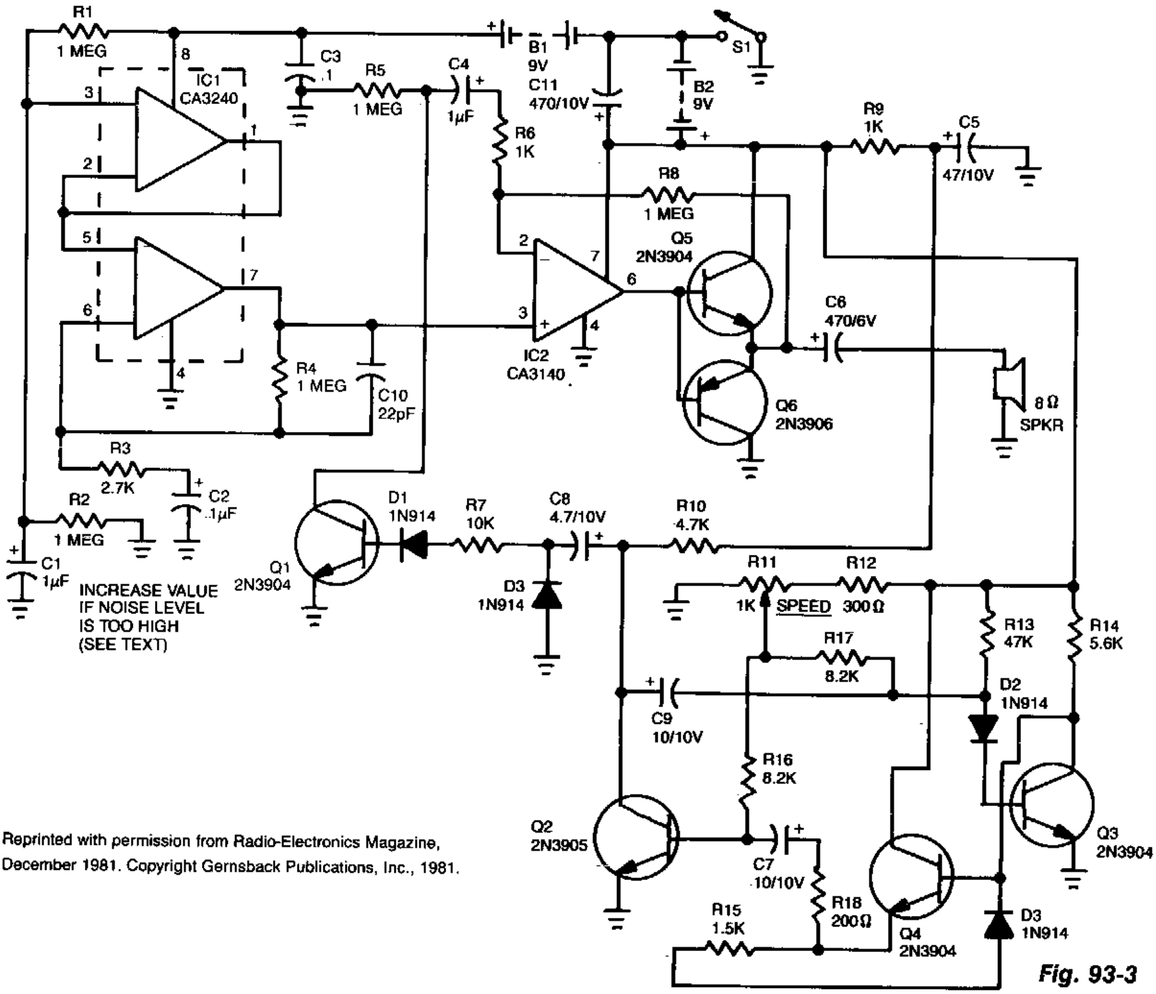


Fig. 93-3

A CA3240 dual MOSFET-input device is used as a white-noise source. Op amp IC2 is used as a driver stage for the push-pull output stage formed by Q5 and Q6. Transistors Q2, Q3, and Q4 form a variable-frequency multivibrator. R11, the speed control, is used to control the multivibrator's frequency. The output is differentiated by C8 and applied to modulator transistor Q1, through D1 and R7. Transistor Q1 modulates the gain of the output amplifier stage by changing the impedance to ground, through R6 and C4. When the multivibrator's frequency is reduced using R11, C8 discharges slowly, creating a sound similar to escaping steam from a stopped locomotive.

To find the proper value for R3, short Q1's collector to ground. Then, increase the value of R3 until the current drain from the power supply is less than 60 mA. Then remove the short from Q1. To see if the device is operating properly, close switch S1 and reduce the resistance of R11. Wait 10 seconds, then rotate R11 slowly. You should hear a sound similar to a steam locomotive picking up speed.

ELECTRONIC BIRD CHIRPER

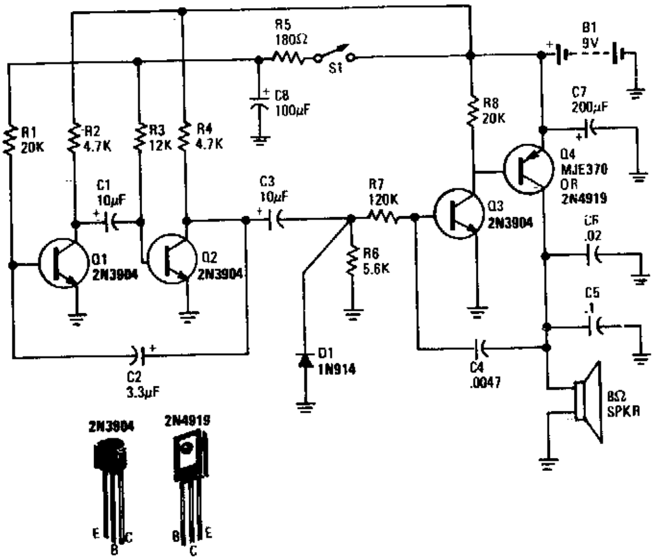
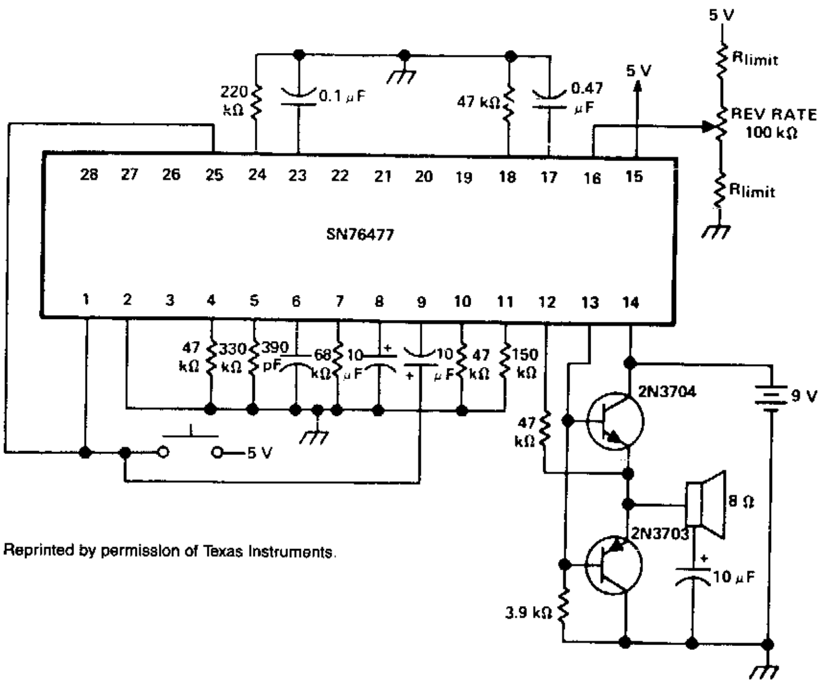


Fig. 93-4

Transistors Q1 and Q2 form the two halves of a free-running multivibrator whose frequency is determined by the voltage across C8. That capacitor is charged and discharged by closing and opening switch S1. Transistors Q3 and Q4 make up a VFO. The output of the free-running multivibrator frequency modulates the Q3/Q4 oscillator, causing the chirping bird sound. The number of chirps per second is determined by the frequency of the Q1/Q2 multivibrator, which also varies. The pitch of the chirps is determined by C5 and C6.

RACE-CAR MOTOR/CRASH



Reprinted by permission of Texas Instruments.

Fig. 93-5

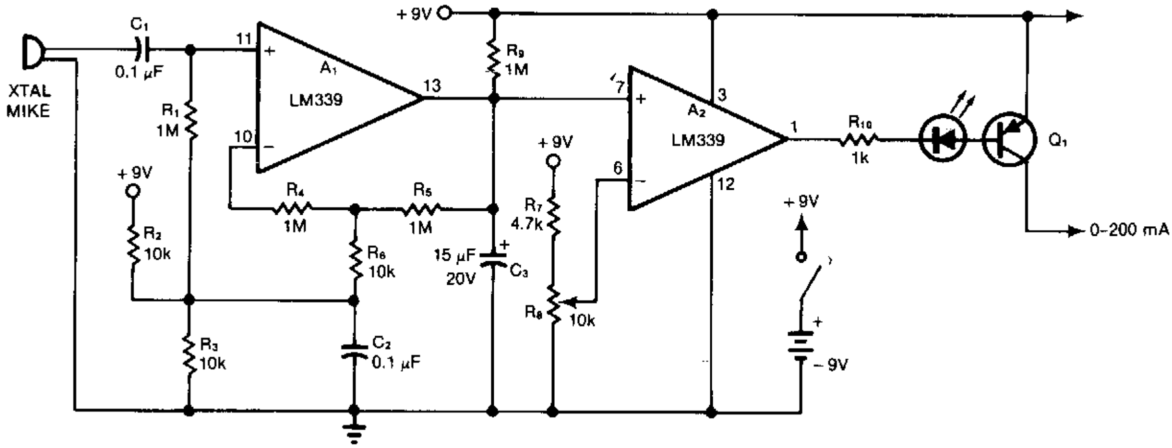
For two simultaneous race-car sounds, the mixer can be multiplexed between the SLF and VCO functions.

94

Sound-Operated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SOUND-ACTIVATED SWITCH

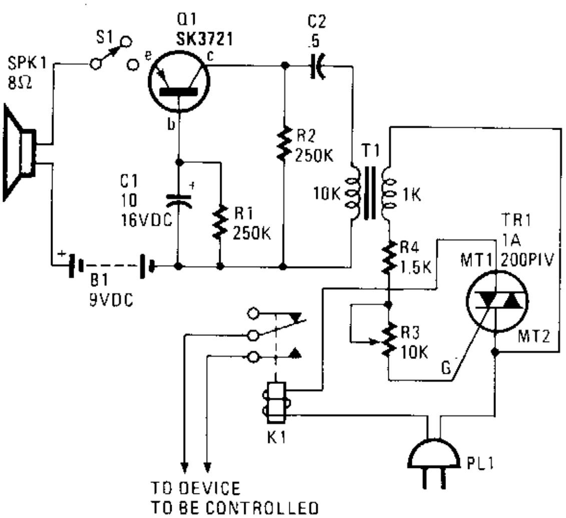


EDN

Fig. 94-1

A1 and A2 are two sections of a quad comparator. The first, A1, functions as an amplifier and detector. Resistors R5 and R6 set the gain at 100; the output of A1 is an open collector to negative-peak-rectify the output with a decay time constant determined by R9 and C3. This dc output is then compared with the reference level selected by R8. A2 triggers switch Q1, and an LED inserted in the base drive of Q1 gives visual indication of switch closure. The standby battery drain is 2 mA. Use potentiometer R8 to select the desired sensitivity.

VOICE-OPERATED SWITCH



The sound picked up by SPKR1, which acts as a microphone, is fed to transistor amplifier Q1. The output of Q1 is applied across coupling transformer T1 and is used to drive the gate circuit of Triac TR1. TR1 is used to lend a latching effect to the action of the relay.

HANDS-ON ELECTRONICS

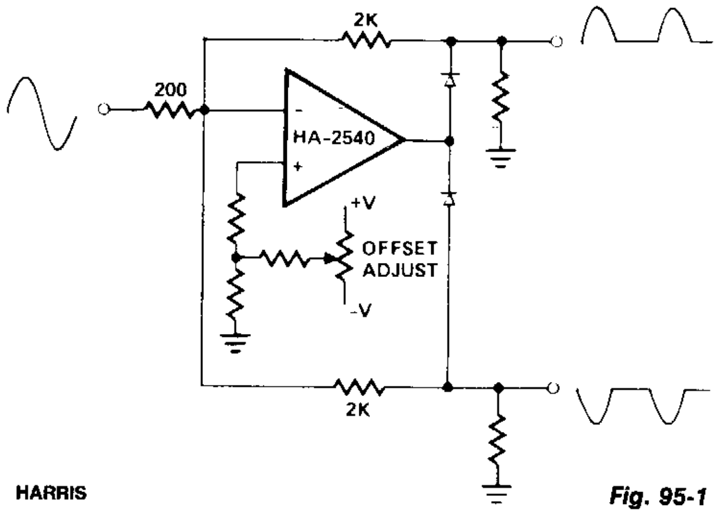
Fig. 94-2

95

Splitters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

WIDEBAND SIGNAL SPLITTER

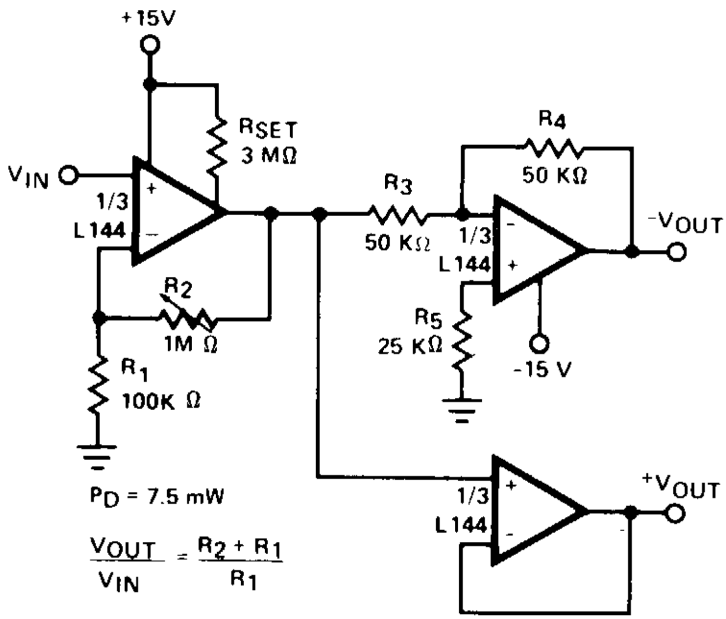


HARRIS

Fig. 95-1

With one HA-2539 or HA-2540 and two low-capacitance switching diodes, signals exceeding 10 MHz can be separated. This circuit is most useful for full-wave rectification, AM detection, or sync generation.

PRECISION PHASE SPLITTER



$$\frac{V_{OUT}}{V_{IN}} = \frac{R_2 + R_1}{R_1}$$

SILICONIX

Fig. 95-2

96

Square-Wave Generators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SQUARE-WAVE PULSE EXTRACTOR

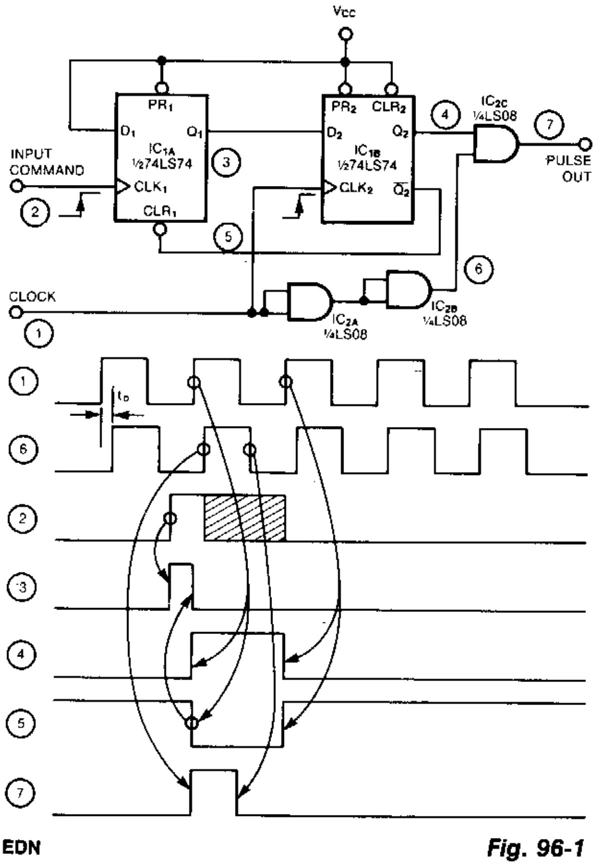


Fig. 96-1

This circuit traps a single positive pulse from a square-wave train. Following the rising edge of an input command, the pulse-out signal emits a replica of one positive pulse of the clock signal simultaneous with the clock signal's next rising edge. The input command signal sets the Q1 output of flip-flop IC1A. Consequently, the next rising edge of the clock signal sets the Q2 output of IC1B, which allows AND gate IC2C to pass the clock signal's next positive pulse. AND gates IC2A and IC2B prevent the generation of brief output glitches by delaying the clock signal by t_D seconds (two propagation delays).

NEARLY 50% DUTY-CYCLE MULTIVIBRATOR

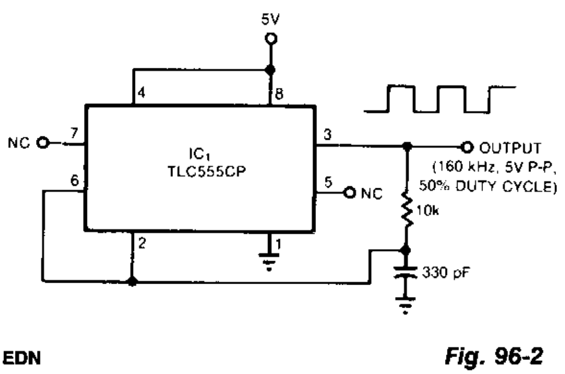
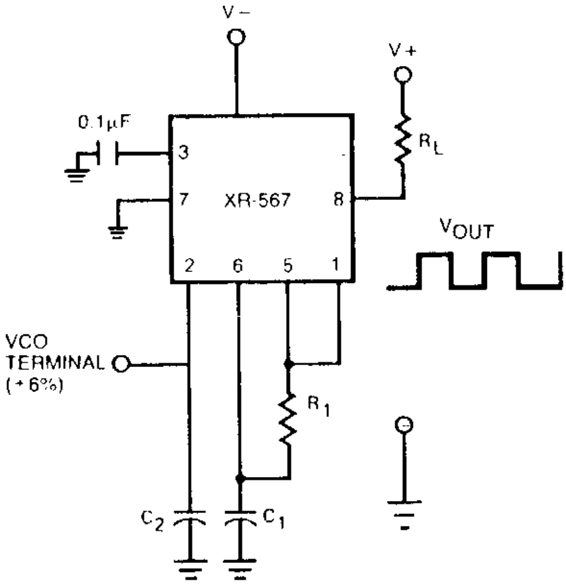


Fig. 96-2

Three factors contribute to the output symmetry. The capacitor charges and discharges through the same external resistor. An internal resistive divider sets accurate switching thresholds within the chip, the bipolar types use dividers, as well. Most importantly, IC1's CMOS output stage switches fully between ground and V_{CC}, avoiding the errors from asymmetry that are often found in a TTL timer's output. The IC's internal switching-threshold tolerances can cause a deviation of several percent from the desired 50% duty cycle. To meet a tighter specification, you might have to select from a group of ICs.

HIGH-CURRENT OSCILLATOR

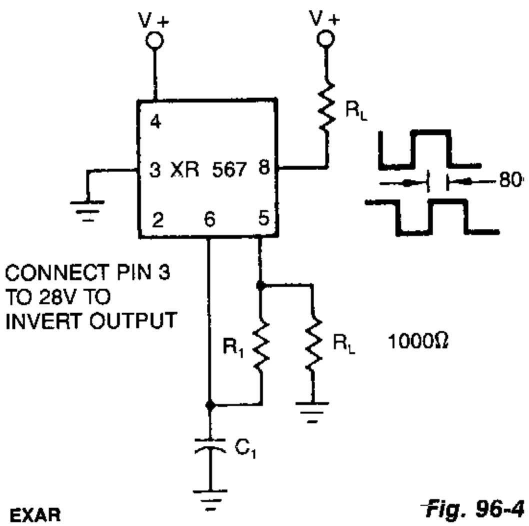


The oscillator output of the XR-567 can be amplified using the output amplifier and high-current logic output available at pin 8. In this manner, the circuit can switch 100-mA load currents without sacrificing oscillator stability. The oscillator frequency can be modulated over ±6% in frequency by applying a control voltage of pin 2.

EXAR

Fig. 96-3

QUADRATURE-OUTPUTS OSCILLATOR



The XR-567 functions as a precision oscillator with two separate square-wave outputs at pins 5 and 8, that are at nearly quadrature phase with each other. Because of the internal biasing arrangement, the actual phase shift between the two outputs is typically 80%.

EXAR

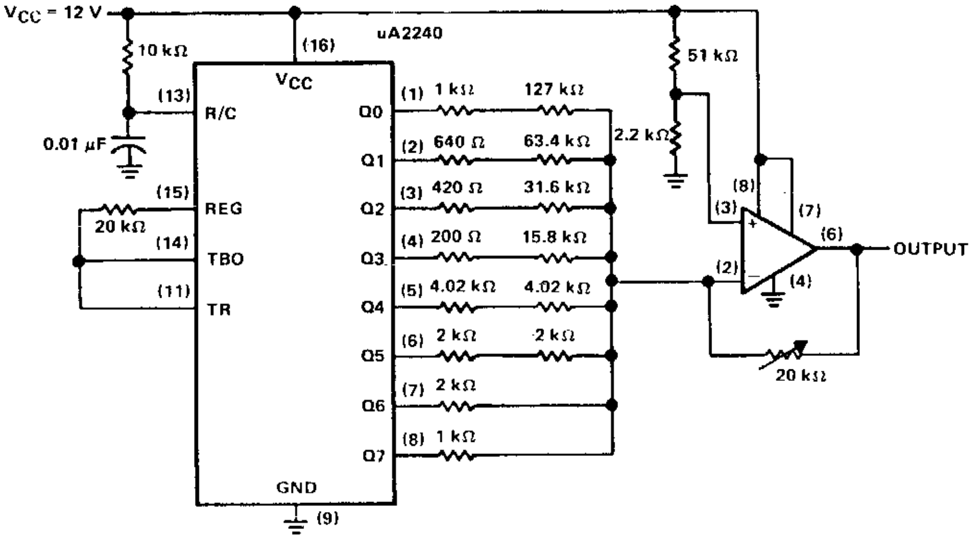
Fig. 96-4

97

Staircase Generators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

UA2240 STAIRCASE GENERATOR



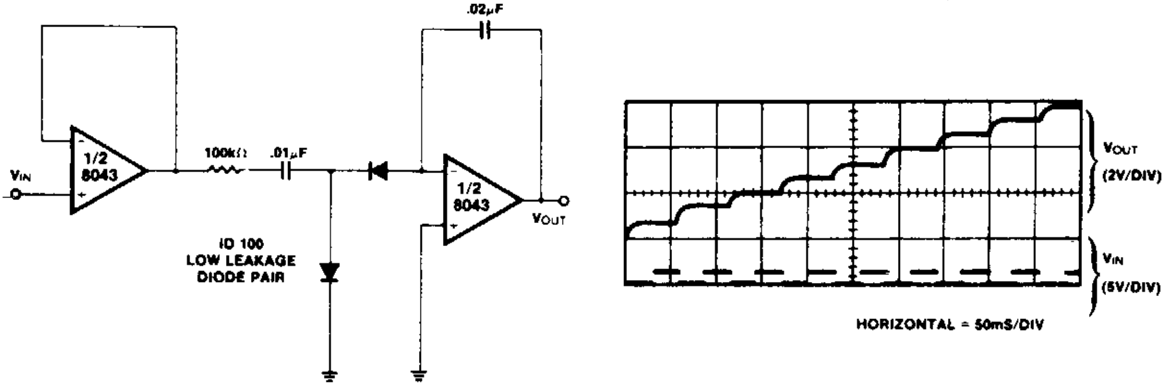
Reprinted by permission of Texas Instruments.

Fig. 97-1

The uA2240 timer/counter, combined with a precision resistor ladder network and an op amp, form the staircase generator. In the astable mode, once a trigger pulse is applied, the uA2240 operates continuously until it receives a reset pulse. The trigger input at pin 11 is tied to the time base output at pin 14, resulting in automatic starting and continuous operation. The frequency of the time-base oscillator, TBO, is set by the time constants R1 and C1 ($f = 1/R1C1$). For this example, a 10-KΩ resistor and a 0.01-μF capacitor form the timing network.

The counter outputs are connected to a precision resistor ladder network with binary-weighted resistors. The current sink through the resistors connected to the counter outputs correspond to the count number. For example, the current sink at Q7, the most significant bit, is 128 times the current sink at Q0, the least significant bit. As the count is generated by the uA2240 eight-bit counter, the current sink through each active binary-weighted resistor decreases the positive output of the op amp in discrete steps. The feedback potentiometer is set at a nominal 10 KΩ to supply a maximum output voltage range. An input of 12 V allows a 10-V output swing. With a 0.5-V input reference on pin 3 of the TCL271, the output will change from 10.46 V maximum, in 256 steps of 38.9 mV per step, to a 0.5 V minimum. Each step has a pulse duration of 100 μs and an amplitude decrease of 38.9 mV. The waveform output is repeated until a reset is applied to the uA2240.

STAIRCASE GENERATOR



INTERSIL

Fig. 97-2

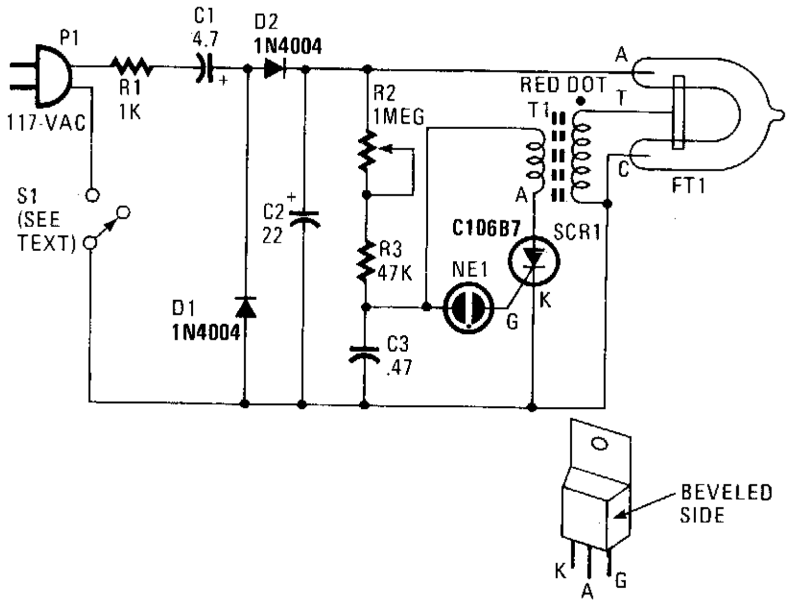
This circuit is a high-input impedance version of the so-called *diode pump* or *staircase generator*. Note that charge transfer takes place at the negative-going edge of the input signal. The most common application for staircase generators is in low-cost counters. By resetting the capacitor when the output reaches a predetermined level, the circuit can be made to count reliably up to a maximum of about 10.

98

Strobe Circuit

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

VARIABLE STROBE LIGHT



HANDS-ON ELECTRONICS

Fig. 98-1

In this strobe-light, two circuits are needed; one circuit charges a capacitor placing 320 Vdc between the cathode and anode of the flashtube. The other circuit provides bursts of approximately 4000 V to trigger the flashtube into conduction. The voltage-doubler works by summing two equal voltages in series, which results in a doubling of the voltage. The 4000 V needed to trigger the flashtube is provided by transformer T1—a voltage step-up transformer that develops 4000 V across its secondary coil when current flows in the primary coil. Silicon-controlled rectifier SCR1 controls the current flow in the primary coil of T1. When SCR1 conducts, current flows suddenly in the primary coil and 4000 Vac spikes appear across the secondary coil. For conduction, SCR1 needs a negative and positive voltage on the cathode and anode, respectively, and a positive voltage on the gate. It is the function of components R2, R3, C3, and NE1 to provide that positive gate voltage and turn on SCR1. Potentiometer R2, resistor R3, and capacitor C3 form an rc timing circuit. Control of charging time of C3 is accomplished by varying that resistance in the circuit. When the voltage on C3 reaches the firing voltage of the neon bulb, it causes NE1 to conduct, thus placing a positive voltage, from C3, on the gate of SCR1. The SCR now turns on and C3 discharges through SCR1 and the primary coil of T1. The 4000 V that is developed across the secondary coil of T1 fires the xenon tube, causing a bright flash. The whole process then repeats itself with C3 charging up, NE1 firing to short out SCR1, and T1 developing 4000 V to trigger the xenon flashtube.

99

Switching Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

RF POWER SWITCH

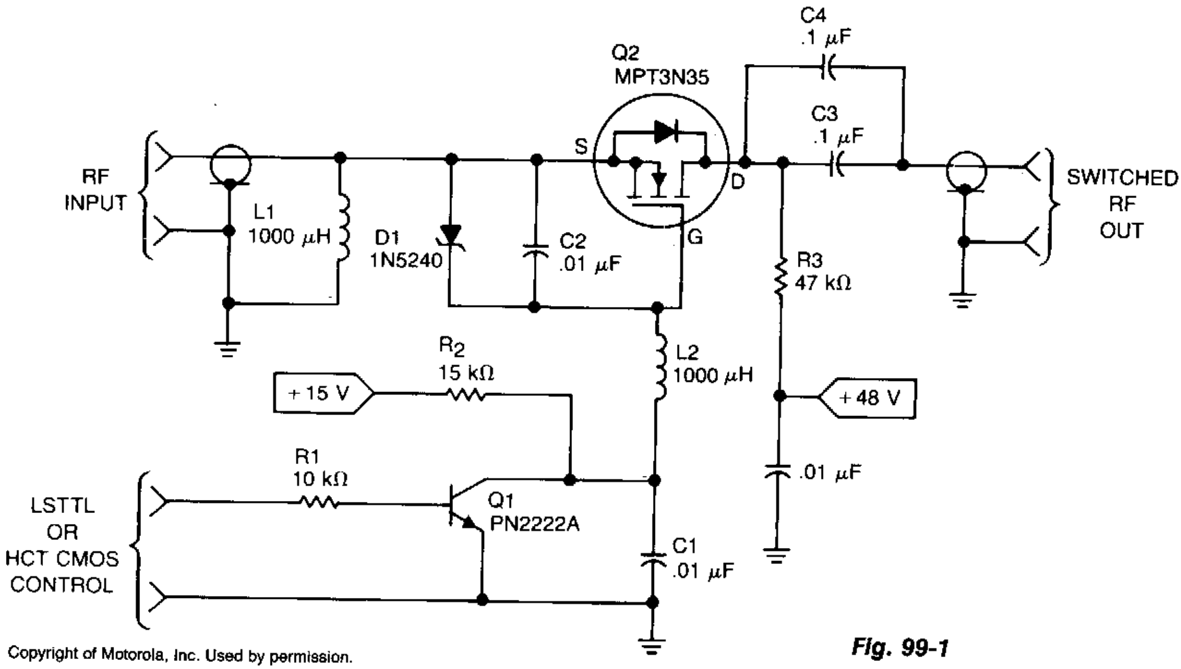
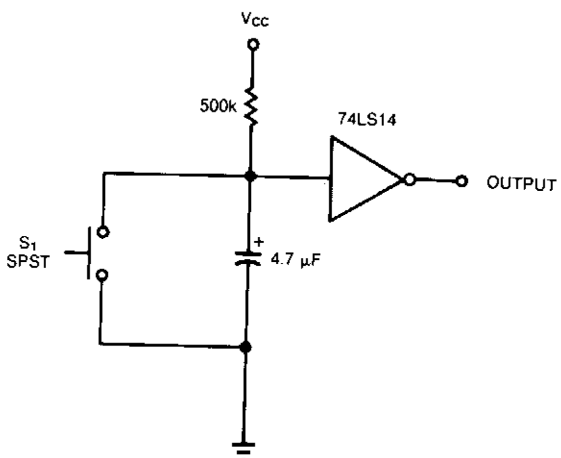


Fig. 99-1

This rf power switch operates at 1.7 MHz with a 50-V source and load. Its on loss is 0.2 dB and its off isolation is 30 dB. It provides 40-W PEP, 45 V_{PEAK} and 0.9 A_{PEAK}. The control input can come from CMOS, TTL, LS, etc., to turn on Q1, which turns on Q2, a TMOS MTP3N35.

SWITCH DEBOUNCER

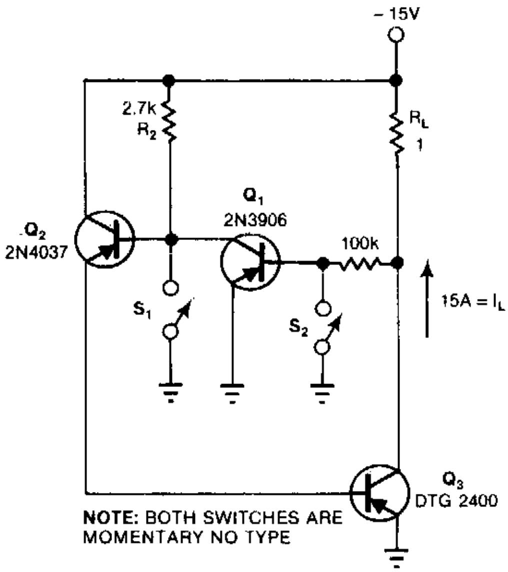


EDN

Fig. 99-2

TTL inverter 74LS14 has an internal 16-KΩ, pull-up resistor that pulls the gate input high when the switch is open. As you close the switch, the 4.7-μF capacitor discharges on the first contact. If the switch contacts bounce open, the internal resistor limits the capacitor's recharge to a rate sufficiently slow to prevent an undesired gate transition before the contacts again close. Note that the circuit correctly debounces the switch for both opening and closing. If you add an external pull-up resistor, you can use a CMOS Schmitt-trigger gate, 74HC14, and a smaller, 0.1-μF, capacitor.

SCR-REPLACING LATCHING SWITCH

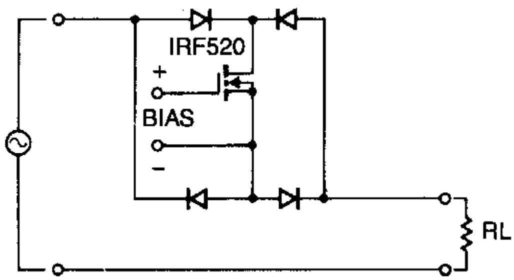


EDN

Fig. 99-3

This circuit provides the turn-on characteristics of an SCR, but turns off with ease. The switch is comprised of three transistors with descending current ratings: Q3 has a high-current rating and Q2 has a medium rating. The current, I1, to be switched is 15 A. Momentarily depressing S2 removes Q1's base drive, turning Q1 off and allowing Q2 to turn on. Q2 then drives the base-emitter junction of Q3, turning Q3 on. Q3's collector-emitter voltage, which serves as Q1's base drive, is essentially zero, keeping Q1 off. To turn Q3 off, depress S1; this action momentarily shunts Q2's base current to ground, reversing the chain of events that turned Q3 on.

ONE-MOSPOWER FET ANALOG SWITCH

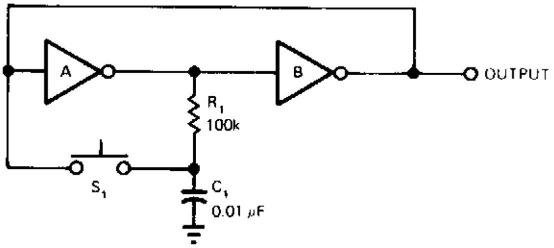


SILICONIX

Fig. 99-4

Using four diode in an array allows using only one MOSPOWER transistor for analog switching. The current flow is controlled by keeping the source-base connection of the MOSFET towards the load. Be sure to use diodes capable of handling the load current and a transistor whose breakdown voltage specification exceeds the peak analog voltage anticipated. Operationally, by increasing the gate-to-source bias voltage, the MOSFET turns on. For applications other than either full-on or full-off, care must be taken not to exceed the dissipation of the MOSPOWER transistor. A suitable heatsink cannot be overstressed in such applications.

ON/OFF INVERTERS



EDN

Fig. 99-5

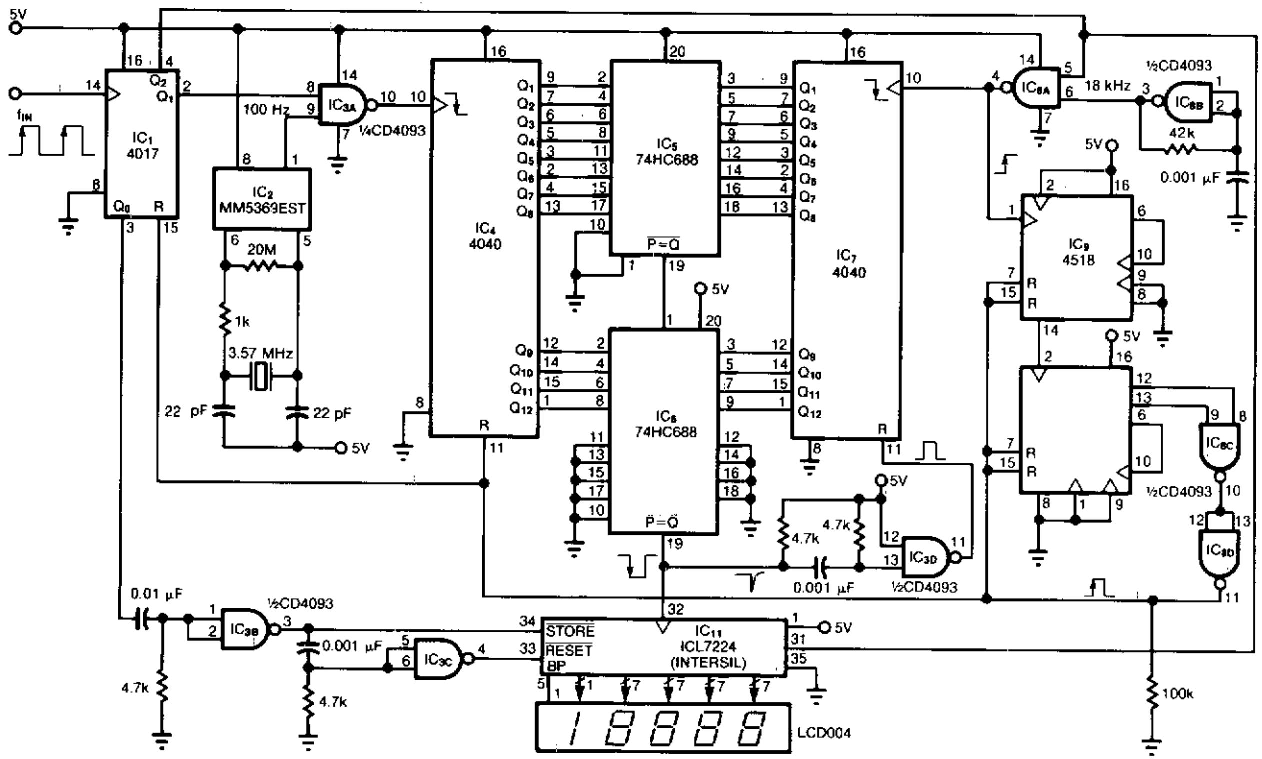
Each time the switch closes, the voltage on C1 causes inverter A to change state, with positive feedback from inverter B. Resistor R1 delays the charging and discharging of C1, making the circuit virtually immune to contact bounce. The circuit works with either CMOS or TTL gates. The values of R1 and C1 are not critical and can be increased for greater contact bounce protection, if needed. Recommended ranges are 10 K to 1 MΩ for R1, and 0.01 to 1.0 µF for C1.

100

Tachometer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

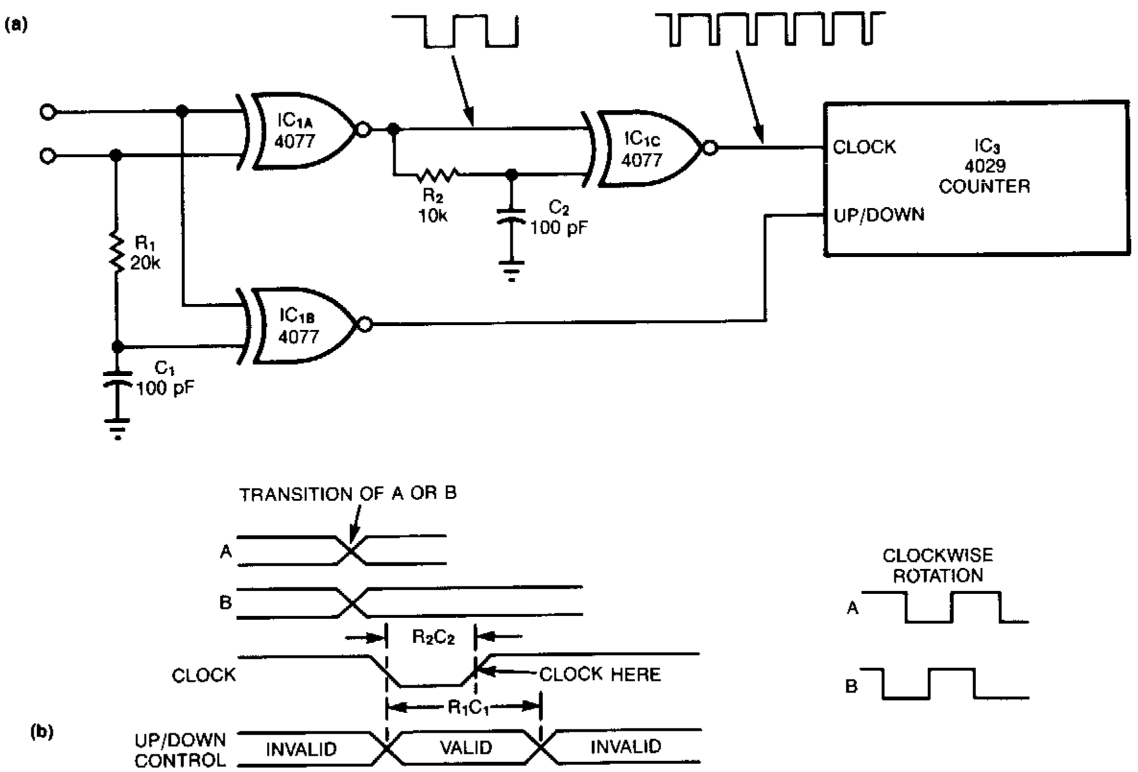
LOW-FREQUENCY TACHOMETER



EDN

Fig. 100-1

This tachometer lets you measure heartbeats, respiratory rates, and other low-frequency events that recur at intervals of 0.33 to 40.96 seconds. The circuit senses the period of f_{IN}, computes the equivalent pulses per minute, and updates the LCD accordingly. Although the decimal readout equals 60 f_{IN}, the circuit doesn't actually produce a frequency of 60 f_{IN}. The computation involves counting and comparison techniques and takes 0.33 seconds.



EDN

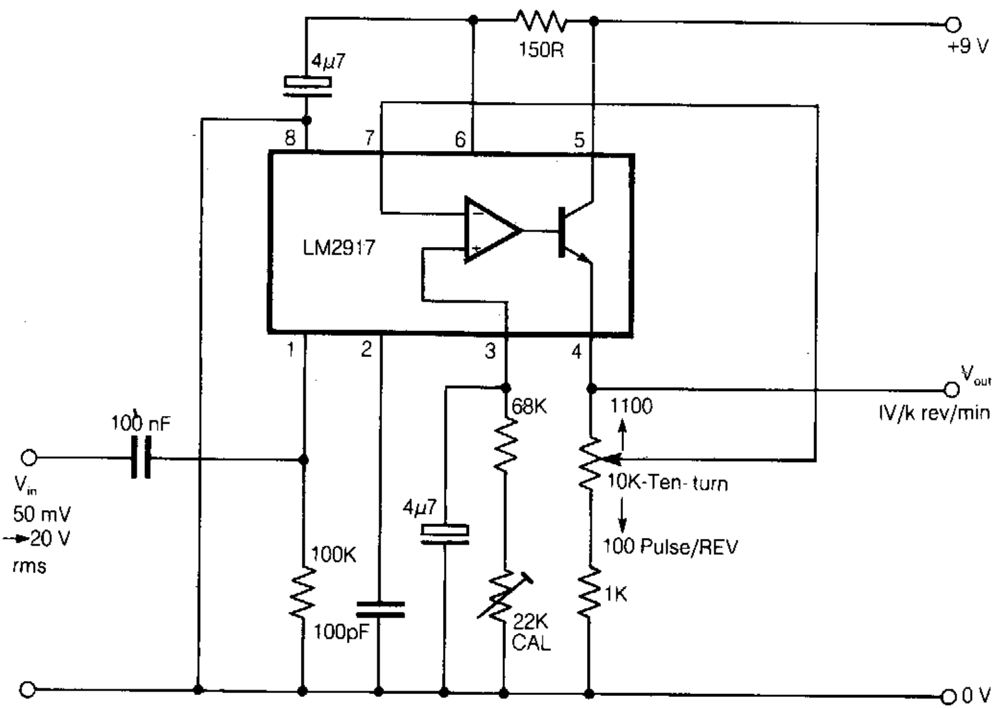
Fig. 100-2

A standard shaft encoder's *A* and *B* ports generate square waves with the same frequency as the shaft turns. The phase of *A* will lead or lag that of *B* by 90°, depending on the direction of rotation. To obtain maximum resolution, the tachometer circuit must count every change of the state for the *A* and *B* signals. Each such change causes a change of state at IC1A's output, followed by a 1-μs negative pulse at the output of IC1C. These clock pulses' positive (trailing) edges cause the counter to count up or down, according to the direction of shaft rotation.

You should set the R1C1 time constant, so that it is approximately twice that of the R2C2 product, to ensure adequate setup and hold times for the up/down signal with respect to the positive clock edges. IC1C supports this timing requirement by producing clock pulses of similar duration for either positive or negative transitions or IC1A.

The exclusive-NOR logic of IC1B generates the correct polarity of the up/down signal when necessary, at the positive clock edges, by combining the *A* value with the *B* value just prior to a transition of *A* or *B*. C1 provides memory by sorting the *B* value voltage for about 2 μs. The maximum frequency for *A* or *B* is approximately $(4R1C1)^{-1}$.

CALIBRATED TACHOMETER



ELECTRONIC ENGINEERING

Fig. 100-3

Here is a simple tachometer circuit for use with a hand-held DVM or portable chart recorder. A novel feature is that the source frequency pulse/rev rate can be directly set on a ten-turn potentiometer to provide a convenient calibration of one V per 1000 rev/min. This is particularly useful when measuring a shaft or engine speed by sensing gear teeth.

The circuit uses an LM2917 IC which is specifically designed for tachometer applications. The ten-turn potentiometer, which provides the pulse/rev setting, is suitably configured in the output amplifier feedback path. The pulse/rev range is 100 to 1100, so the potentiometer dial mechanism should be set to start at 100 to provide direct calibration.

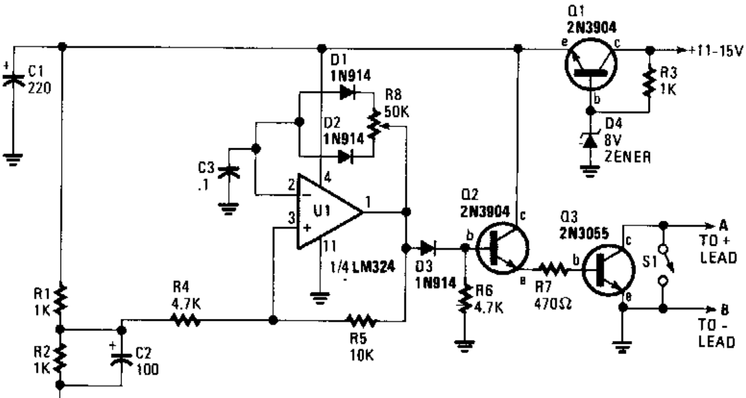
The IC's internal 7.5-V zener provides stable operation from a 9-V battery. The tachometer accepts an input signal between 50 mv and 20 V rms and has an upper speed limit of 6000 rev/min with the component values shown.

101

Tape-Recorder Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

EXTENDED-PLAY CIRCUIT



HANDS-ON ELECTRONICS

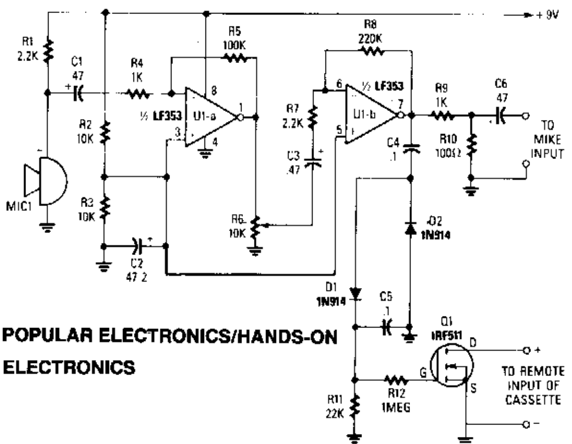
Fig. 101-1

A single op amp—one of four contained in the popular LM324—is operating in a variable pulsewidth, free-running squarewave oscillator circuit, with its timed output driving two transistors that control the on/off cycle of the tape-drive motor.

The oscillator's positive feedback path holds the secret to the successful operation of the variable on/off timing signal. The two diodes and pulsewidth potentiometer R8 allows the setting of the on and off time, without affecting the oscillator's operating frequency. One diode allows only the discharge current to flow through it and the section of R8 that it's connected to. The other diode, and its portion of R8, sets the charge time for the timing capacitor, C3. Since the recorder's speed is controlled by the precise off/on timing of the oscillator, a simple voltage-regulator circuit (Q1, R3, and D4) is included.

Connecting the speed control to most cassette recorders is a simple matter of digging into the recorder and disconnecting either of the motor's power leads, the ground or common side might be best, and connecting the recorder through a length of small, shielded cable to the control circuit. In some recorders, a remote input jack is furnished to remotely turn on and off the recorder. Before going in and modifying a recorder with a remote jack, try connecting the circuit to the external remote input.

SOUND-ACTIVATED SWITCH

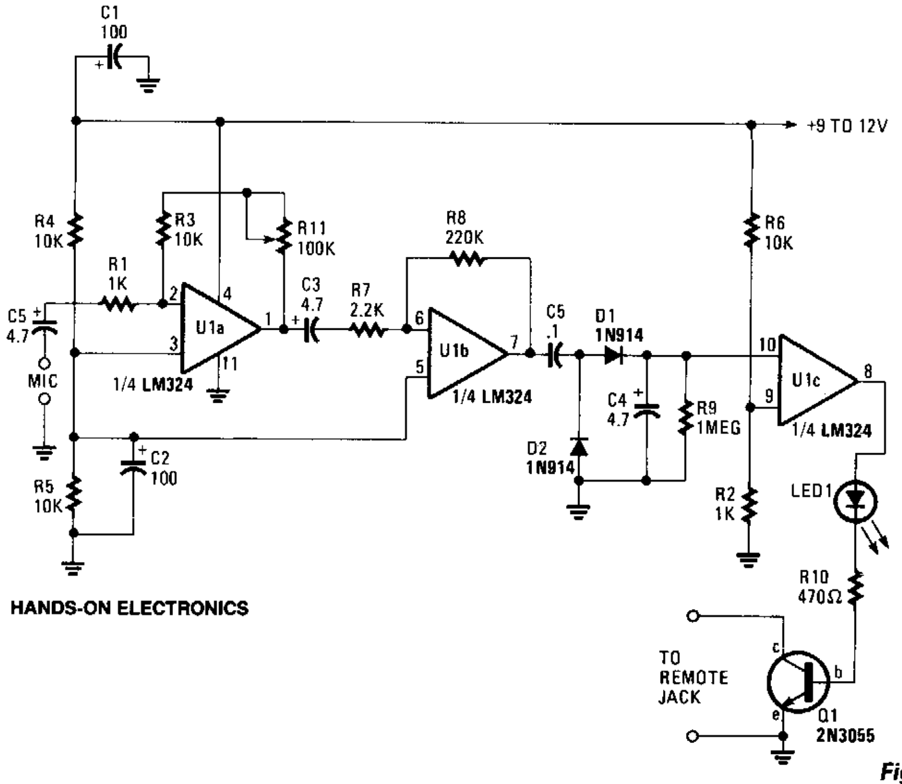


POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

A sensitive electret microphone picks up the sound and feeds the signal to a two-stage amplifier circuit, consisting of U1a and U1b. The amplified output of U1b is fed to a voltage-doubler circuit (comprised of D1, D2, C4, and C5). The output of the doubler is input to the gate of Q1. When the dc voltage reaches the gate's threshold level, Q1 switches on, starting the recorder. Resistor R6 sets the circuit's sensitivity and should be experimented with to obtain the optimum adjustment.

Fig. 101-2

SOUND-ACTIVATED TAPE SWITCH



HANDS-ON ELECTRONICS

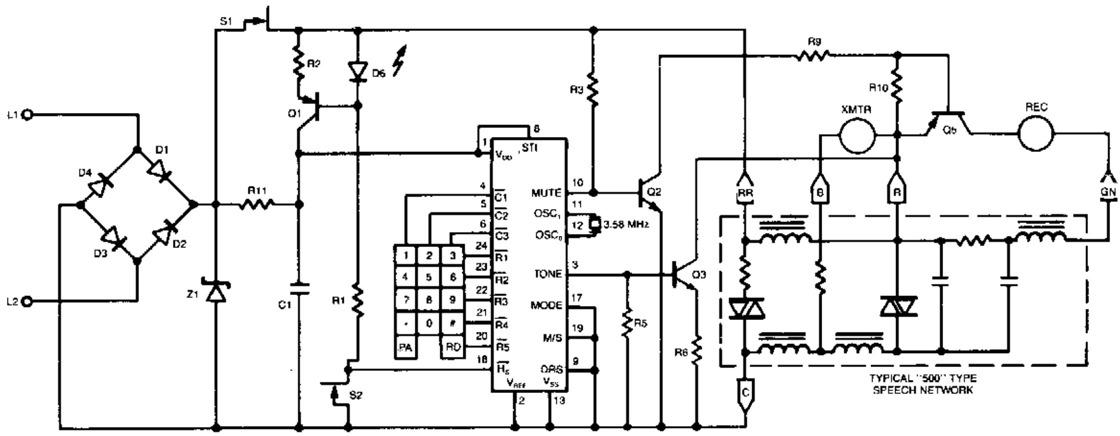
Fig. 101-3

This circuit can cause a cassette recorder to automatically turn on and record when a sound or noise is present. Another use, is when the sound-activated switch is used to turn on a cassette player so that it operates as a burglar-alarm detector and sounder. Op amps U1a and U1b are connected in tandem to amplify the sounds picked up by the detector's mike. The amplified audio voltage, output at pin 7 of U1b, is fed to a voltage-doubler circuit, consisting of D1 and D2. The elevated voltage from the doubler circuit is input to the positive input of op amp U1c, which is operating as a simple comparator circuit. The other input of U1c is connected to a voltage divider that sets the switching point for the dc signal voltage, to turn on when the signal level is greater than about 1.5 V. As the comparator switches on, its output at pin 8 becomes positive and supplies a forward bias to turn on D3 and Q1, which in turn, starts the recorder. The rc combination of C4/R9 sets the cassette's run time after the input sound has ceased, preventing the recorder from chopping-up or turning-off between closely spaced sounds or words picked up by the mike. The delay time is roughly 6 to 8 seconds. R11 sets the circuit's gain. Connect a low-impedance cassette mike to the amplifier's input, and connect the output of Q1 to the cassette's remote input or to the internal input and set the recorder to the record position. Talk and adjust the amplifier's gain with R11 for the desired sensitivity.

102

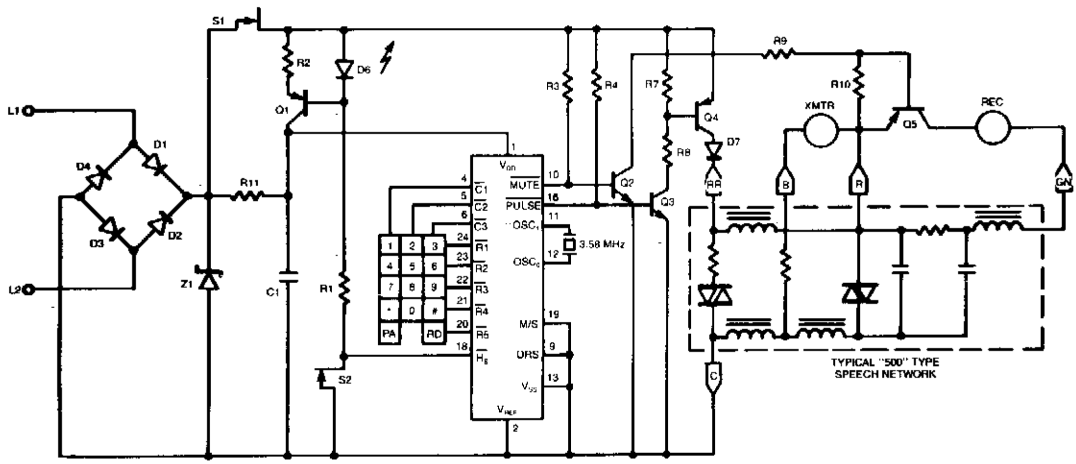
Telephone-Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.



R1 = 560KΩ	R9 = 3KΩ	Q1 = 2N5401	D1 = 1N4004	C1 = 68μF
R2 = 14Ω	R10 = 100KΩ	Q2 = 2N5550	D2 = 1N4004	S1 = HOOK SWITCH
R3 = 470KΩ	R11 = 20KΩ	Q3 = 2N5550	D3 = 1N4004	S2 = HOOK SWITCH
R5 = 330Ω		Q5 = 2N5401	D4 = 1N4004	Z1 = 120V 1 WATT ZENER
R6 = 120KΩ			D6 = LED	

Typical Tone Dialing Application Circuit



R1 = 560KΩ	R7 = 100KΩ	Q1 = 2N5401	D1 = 1N4004	C1 = 68μF
R2 = 14KΩ	R8 = 3KΩ	Q2 = 2N5550	D2 = 1N4004	S1 = HOOK SWITCH
R3 = 470KΩ	R10 = 100KΩ	Q4 = 2N5401	D4 = 1N4004	Z1 = 120V 1 WATT ZENER
R4 = 330Ω		R11 = 20KΩ	Q5 = 1N4004	D6 = LED
			= 2N5401	D7 = 1N4004

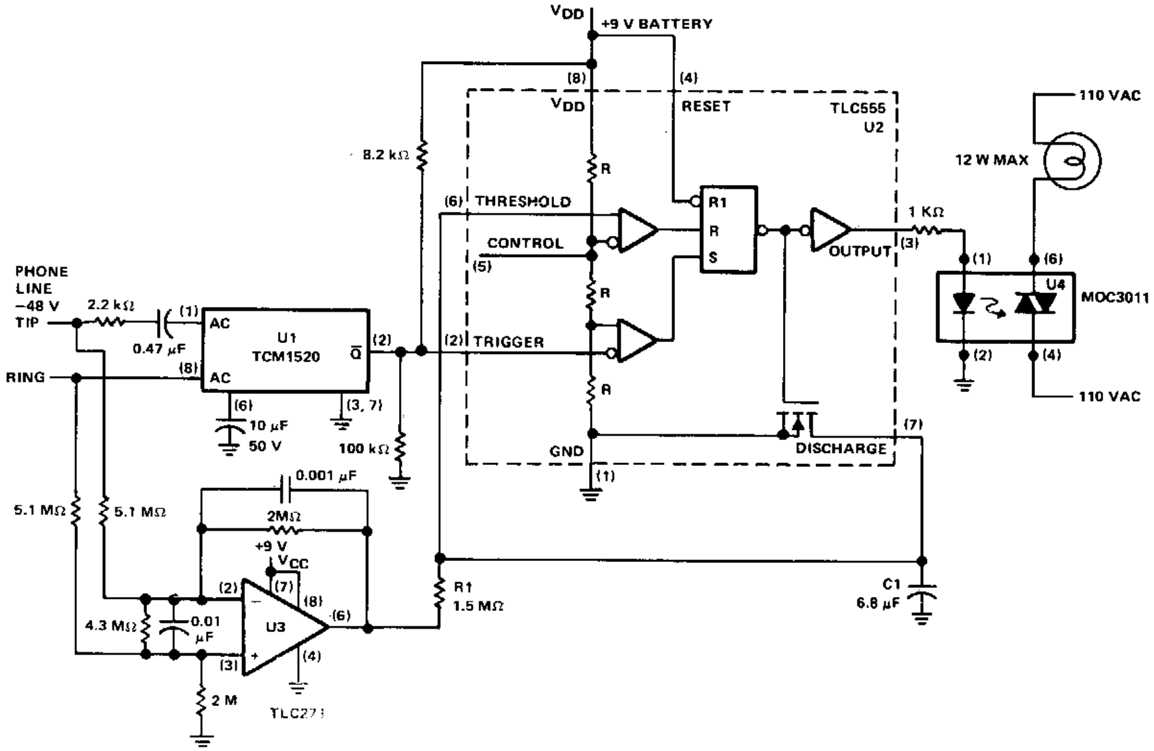
Typical Pulse Dialing Application Circuit

EXAR

Fig. 102-1

The XR-T5990 single-chip pulse/tone Dialer is a silicon gate CMOS circuit which performs both pulse and tone functions. It is designed to operate directly from the telephone line or on a separate small power supply. A 17-digit buffer is provided for redial feature.

TELEPHONE-CONTROLLED NIGHT LIGHT

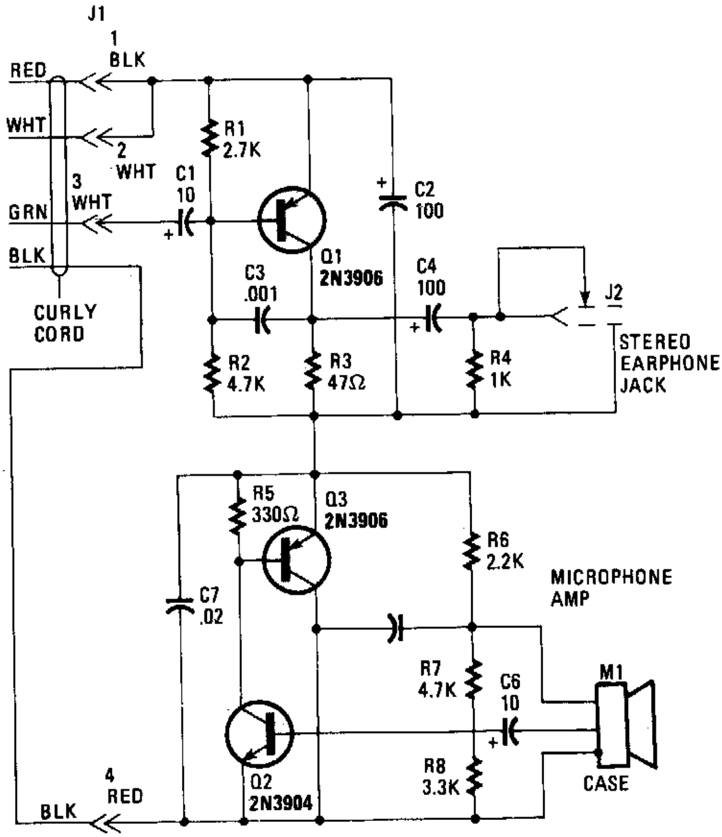


Reprinted by permission of Texas Instruments.

Fig. 102-2

When the telephone rings, or when the handset is lifted, the night light is turned on and remains on while the conversation takes place. When the handset is replaced in the cradle, the light remains on for about 11 s. During standby conditions, the −28 Vdc bias on the phone line maintains the output of U3 in a high state. When the ac ring signal is applied to the phone line, it is processed by the ring detector U1, producing a negative output pulse at pin 2 for each ring. These pulses trigger U2, causing its output to become high and the discharge transistor to turn off. The high output of U2 activates optoisolator U4, which turns on the night light. Each ring retriggers the timer and discharges C1, preventing it from reaching the $^2/_3$ V_{DD} threshold level. Thus, the night light will remain on while the phone is ringing and for about 11 s after the last ring. After 11 s, C1 will be charged to the U2 threshold level ($^2/_3$ V_{DD}) resulting in the U2 output returning to a low level and its discharge output turning on, discharging C1. The lamp will turn off if the phone is not answered.

When the phone is answered, a 1-KΩ load is placed across the phone. This removes the differential input to op amp U3, causing its output to become low, and capacitor C1 starts discharging through R1. As long as the voltage across C1 remains low, timer U2 cannot start its cycle and the lamp will remain on. When the phone is hung up, the low impedance is removed from the phone line and the differential voltage across the line causes the U3 output to become high. This allows C1 to start charging, initiating the timing that will turn off the night light.



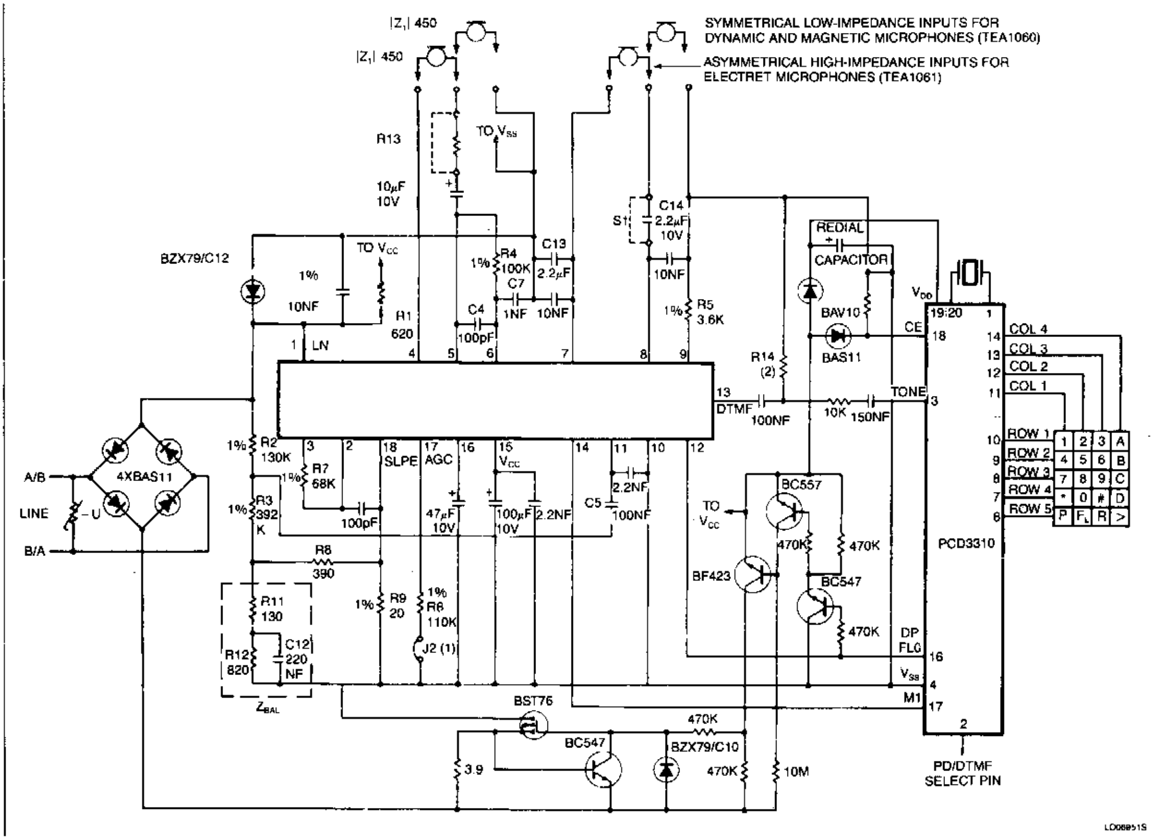
HANDS-ON ELECTRONICS

Fig. 102-3

Transistor Q1 of the headset amplifier circuit amplifies the 30 mV signal, that would have gone to the earphones, to .5 V, which sufficiently drives the stereo earphones. Capacitor C1 blocks any dc current from shorting back into the telephone base. Capacitor C2 provides the very important ac signal short around the amplifier. Capacitor C3 provides high-frequency rolloff characteristics and prevents the amplifier from oscillating. Capacitor C4 is a dc block to the 35-Ω impedance of the stereo earphones, and resistor R4 bleeds off any charge build up to prevent a popping sound when the stereo earphones are plugged into the mini-earphone jack J2. The headset amplifier has only about 2 Vdc across it. The microphone amplifier circuit is composed of transistors Q2 and Q3 in an inverted-Darlington configuration.

Another, and perhaps easier, way to understand the operation of this circuit is to consider Q3 as an emitter-follower stage. The electret microphone has a built-in FET IC amplifier that needs at least 3 V at 0.4 mA of clean supply power in order to provide an output impedance of 200 to 800 Ω. Resistors R6 and C5 provide that clean dc power to the FET IC and also provide the bias to Q2 without an ac feedback, which would have reduced Q2's gain. Capacitor C6 blocks the output dc bias from the FET IC.

ELECTRONIC TELEPHONE SET WITH REDIAL

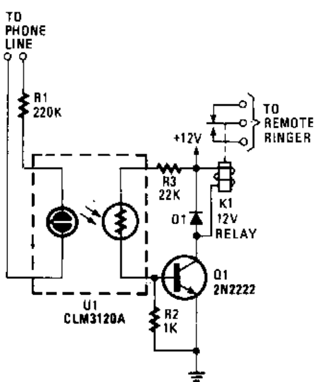


NOTES:

1. Automatic line compenstion obtained by connecting R6 to V_{SS}.
2. The value of resistor R14 is determined by the required level at LN and the DTMF gain of the TEA1060.

SIGNETICS *Fig. 102-4*

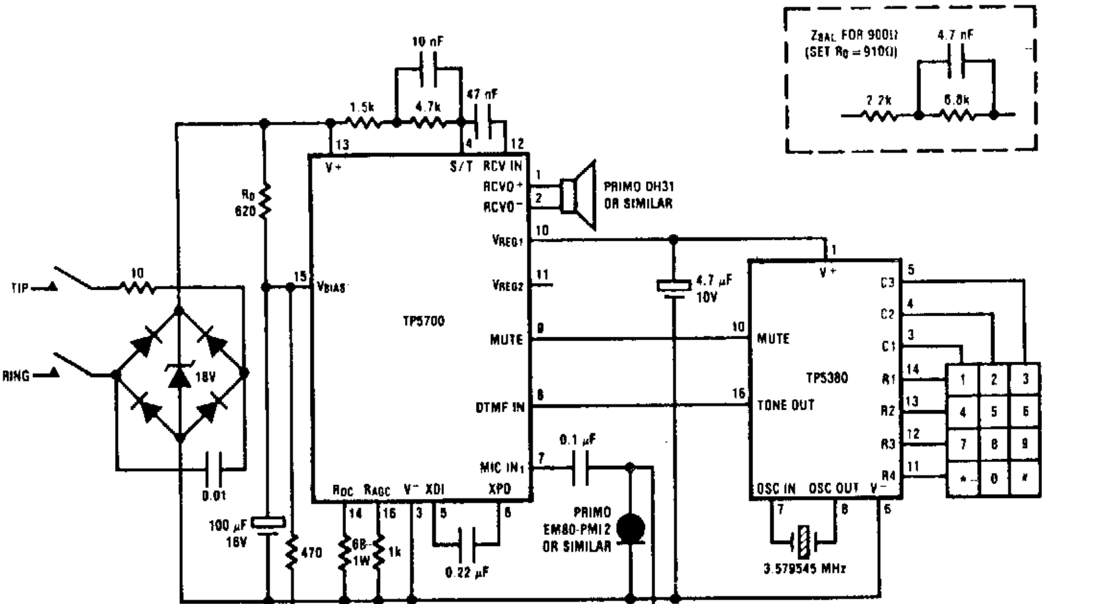
RINGER RELAY



When the phone rings, the ring signal from the telephone company lights a neon lamp within a CLM3120 optocoupler. That causes a drop in the resistance of the CdS cell output of the device, turning on transistor Q1. When Q1 turns on, relay K1 is energized. The circuit should be connected in series with the lamp that is to be activated.

POPULAR ELECTRONICS *Fig. 102-5*

TONE-DIALING TELEPHONE



NATIONAL SEMICONDUCTOR CORP.

Fig. 102-6

This circuit shows the TP5700 directly interfacing to a low-voltage DTMF generator. V_{REG1} supplies the necessary 2-V minimum bias to enable the TP5380 to sense key closures and pull its mute output high. V_{REG1} then switches to a 3-V regulated output to sustain the tone dialer during tone generation. The TP5700 DTMF input incorporates the necessary load resistor to V− and provides gain, plus AGC action, to compensate for loop length. A muted tone level is heard in the receiver. For DTMF generators with a higher output level than the TP5380, a resistive potentiometer should be added to reduce the level at the speech circuit DTMF input.

TELEPHONE REPEATER

POPULAR ELECTRONICS

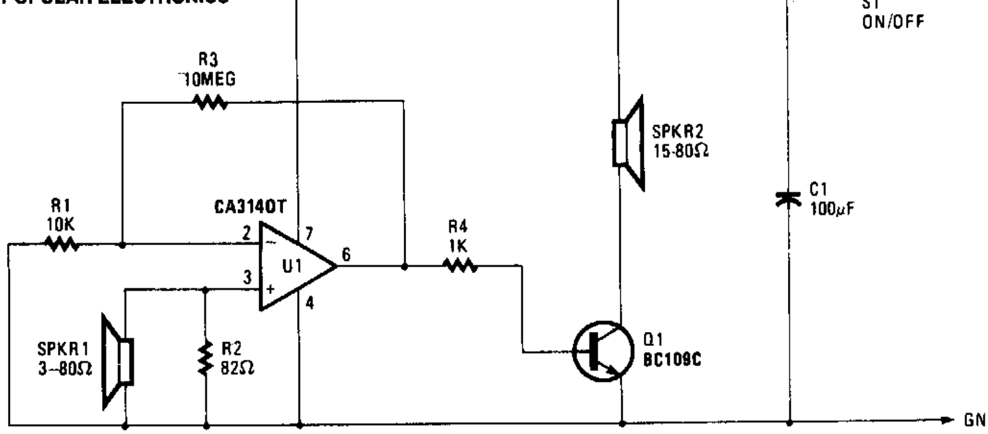
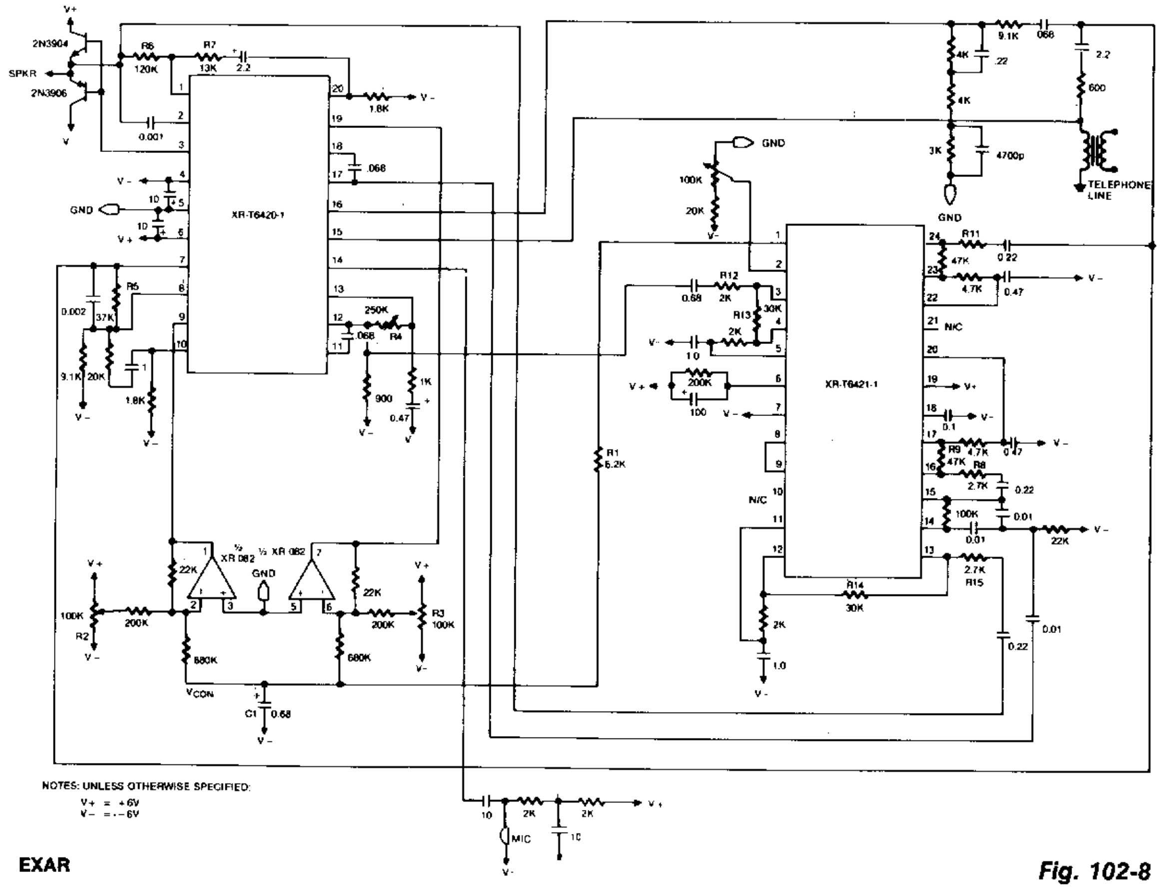


Fig. 102-7

SPEAKERPHONE

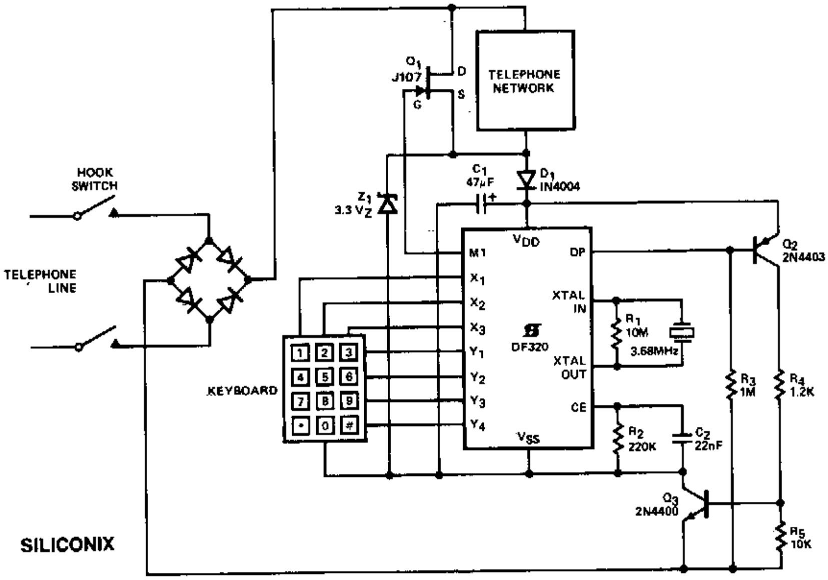


EXAR

Fig. 102-8

This circuit consists of two audio channels, a control circuit, and a hybrid interface circuit. The gain of each audio channel is controlled by the control circuitry, with the use of a voltage controlled amplifier (VCA). The inputs to the control circuit are obtained from each of the audio channels. The hybrid interface circuit performs three important functions. First, it couples the T_X channel signal to the telephone line. Second, it couples the signal on the telephone line to the R_X channels. And, finally, it cancels a majority of the T_X signal that can couple into the R_X channel. The amount of T_X signal that appears on the R_X channel is called sidetone.

SERIES TELEPHONE CONNECTION

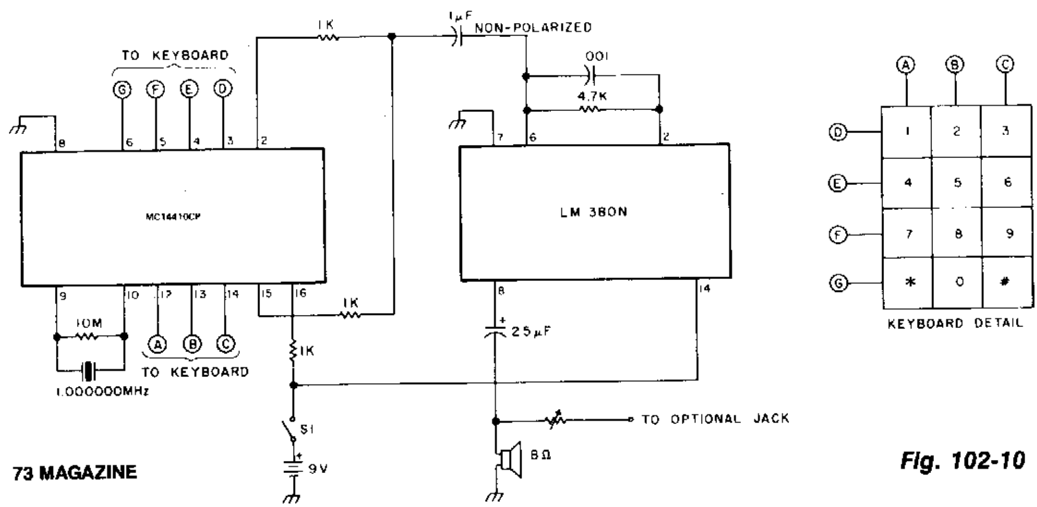


SILICONIX

Fig. 102-9

Here is a simple method of series connection into the telephone set suitable for PABX or short line applications. When the telephone handset is lifted, C1 is charged via D1 to (V_{Z1} − 0.7) V and DF320 power on reset occurs. When the first keyed digit is recognized, M1 goes to logic *1*, muting the telephone network by switching on the low on resistance JFET Q1, and maximizing the line-loop current for impulsing. Impulsing occurs through DP switching Q2, and hence Q3 turns off. Rapid discharge of C1 through Z1 is prevented during line break by blocking diode D1. When dialing is complete, the circuit returns to the static standby condition, and Q1 is switched off. The circuit reset, during a line interruption by the cradle switch, is for the parallel connection mode.

SIMPLE TOUCHTONE™ GENERATOR



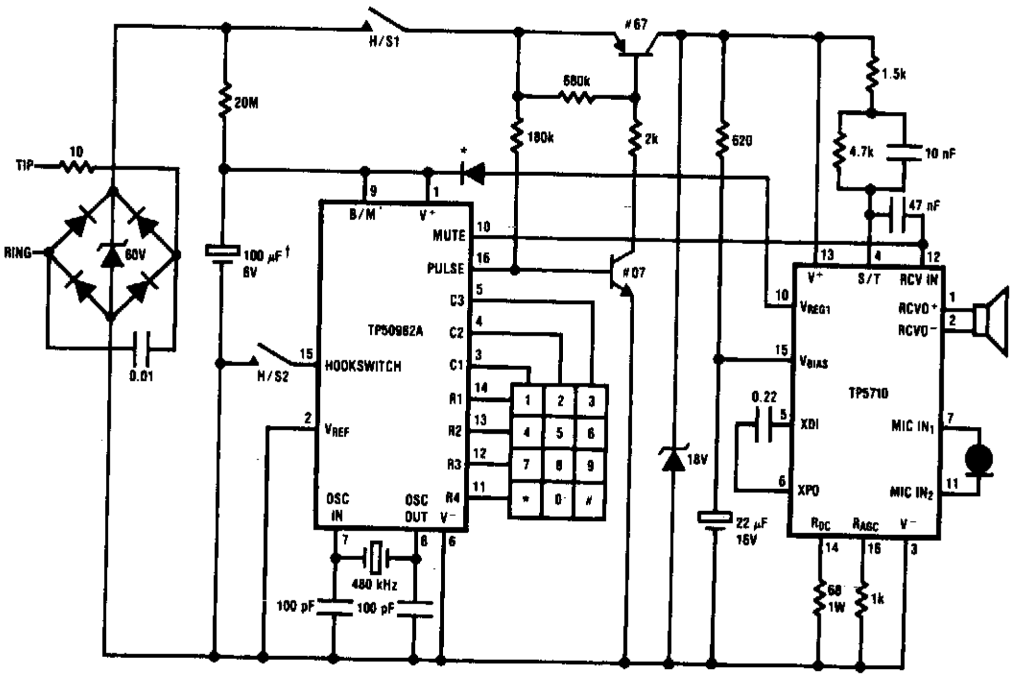
73 MAGAZINE

Fig. 102-10

SIMPLE TOUCHTONE™ GENERATOR (*Cont.*)

The oscillator is a Motorola MC14410CP chip using a 1-MHz crystal. The chip generates both the high and low tones, feeding the energy to the amplifier through 1-K resistors and the 1-μF capacitor. Values for the output resistors can vary from a few hundred Ω to about 60 KΩ. The value of the resistor shunting the crystal can vary from about 3 to 15 MΩ. The amplifier consists of an LM-380N.

PULSE-DIALING TELEPHONE

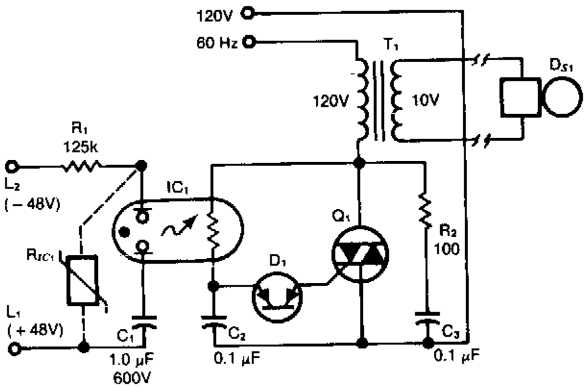


* Select as necessary to suit mic sensitivity
† Low leakage type

NATIONAL SEMICONDUCTOR CORP. *Fig. 102-11*

The TP5700 or TP5710 can reduce the number of components required to build a pulse-dialing telephone, as shown. The usual current source can be eliminated by using the V_{REG1} output to power a TP50982A low-voltage (1.7 V) pulse dialer via a blocking diode. A low forward-voltage drop diode such as a Schottky type is necessary because V_{REG1} is used in its nonregulated mode and its output voltage might fall to 2 V on a 20-mA loop. A 100-μF decoupling capacitor is required to hold up the pulse dialer supply voltage during dialing. This capacitor will take about one second to charge up when the telephone is first connected to the line, but thereafter, the 20-MΩ resistor, required to retain the last-number dialed memory, will keep this capacitor charged. Partial muting is obtained by directly connecting the N-channel open-drain mute output of the pulse dialer to the RCV in pin on the speech circuit. A fully muted pulse dialer design requires the use of a shunt-mode dialer, such as the TP50981A or TP50985A.

OPTICALLY INTERFACED RING DETECTOR

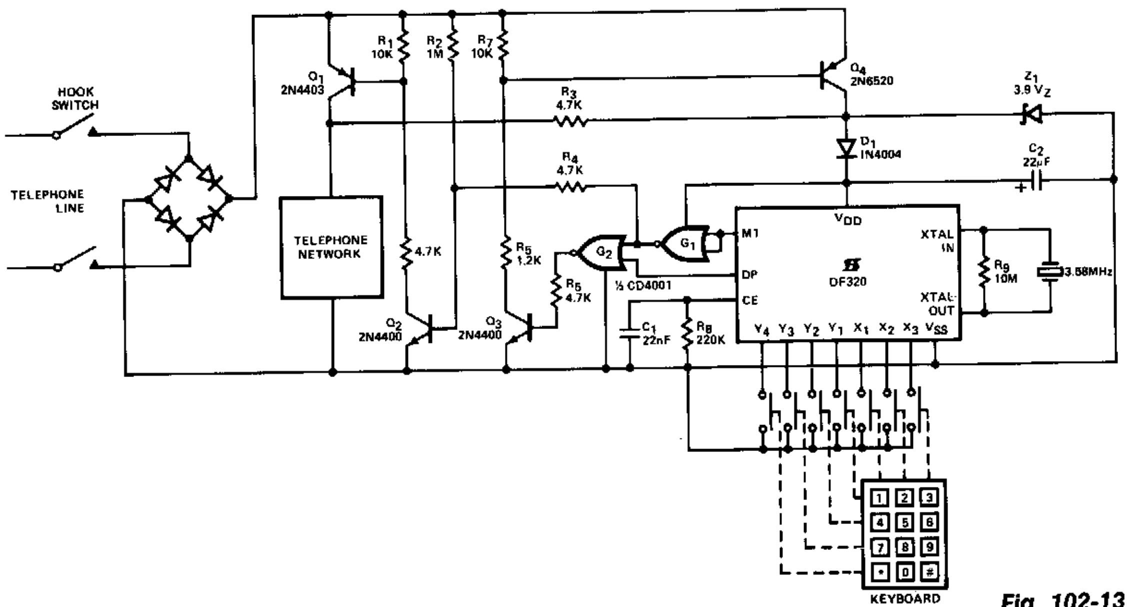


This ring detector, utilizing a neon-LDR (light-dependent resistor) optocoupler, simplifies interfacing with telephone lines.

EDN

Fig. 102-12

PARALLEL TELEPHONE CONNECTION

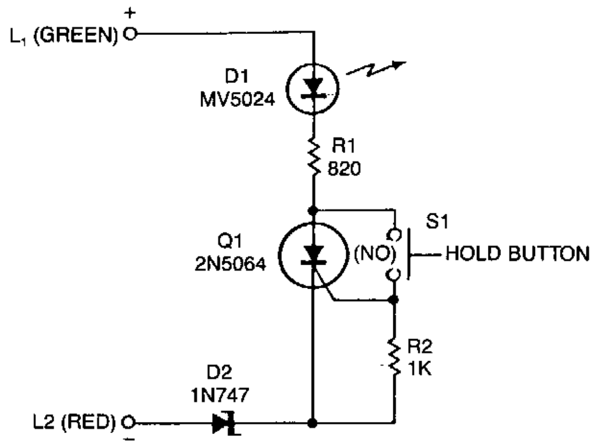


SILICONIX

Fig. 102-13

When the handset is lifted and power is applied to the circuit, Q2 is fed base current through R2, which in turn drives Q1. C2 is charged via R3 in series with D1 to (V_{Z1} − 0.7) V. When the minimum operating V_{DD} voltage is reached, power on reset occurs via the rc network of C1 and R8. Q2 is maintained in the on condition by G1, while Q3, and hence Q4, are held off by G2. The DF320 network appears in parallel with the telephone as an impedance more than 10 KΩ in the standby condition with the telephone network connected in circuit through Q1. On recognition of the first keyed digit, the DF320 clock is started. M1 then goes to logic 1 causing Q2 and Q1 to turn off, and Q3 and Q4 to turn on. Hence, the majority of the line loop current now flows through Q4 and Z1. When impulsing occurs, Q3 and Q4 are turned off by DP acting on G2. Line loop current is then reduced to approximately 50 μA taken through R2, R4, and G2 in series. When dialing in, complete M1 goes to logic *O*, causing the telephone network to be reconnected. The DF320 then returns to the static standby condition. If the line loop is interrupted by the cradle switch during dialing, impulsing will continue until C2 discharges to a voltage, such that R8 pulls CE to logic *O*, causing the DF320 to reset. The diode bridge protects the network from line polarity reversal.

ADD-ON TELEPHONE HOLD BUTTON



EDN

Fig. 102-14

A sensitive-gate SCR provides a line-holding current of 20 to 40 mA, depending on loop resistance. It also lights an LED to give the user a positive indication that the telephone line is on hold. The 20 to 40 mA should prove sufficient to hold the majority of lines, but it might require increasing—by decreasing the size of R1—in individual instances. When any receiver in the same loop is lifted, the low impedance of the off-hook telephone set shunts holding current away from the SCR, thereby releasing the line and extinguishing the LED. Zener diode D2 ensures that the line-holding current drops below the SCR's minimum conduction current. If the calling party tires of waiting on hold and hangs up, the release of the central-office relays from the calling side also releases the line from the hold mode.

TELEPHONE HANDSET ENCODER

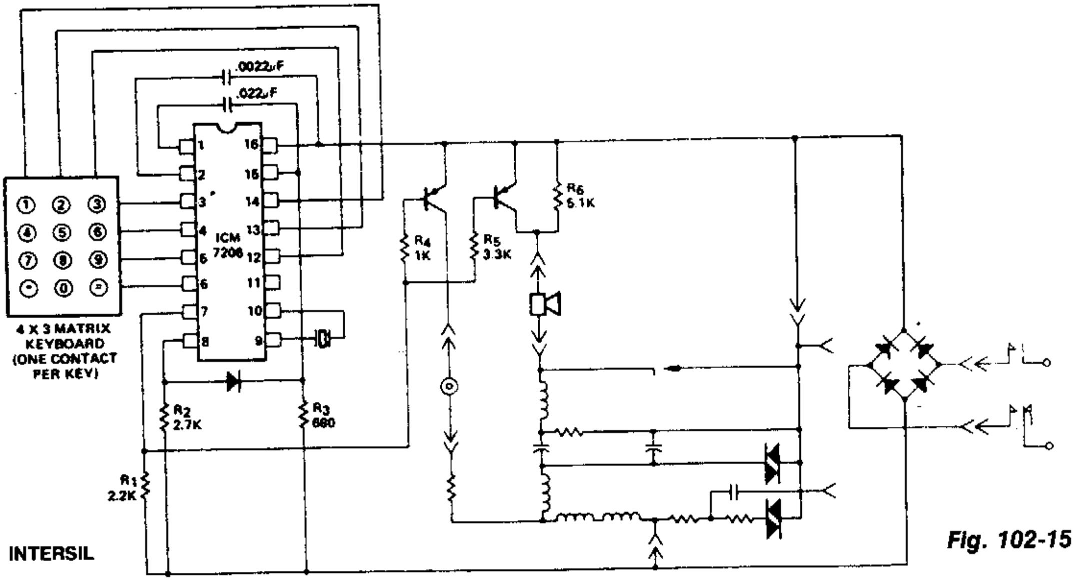


Fig. 102-15

This encoder uses a single contact per key keyboard and provides all other switching functions electronically. The diode connected between terminals 8 and 15 prevents the output from going more than 1 V negative with respect to the negative supply V_{SS}. The circuit operates over the supply voltage range from 3.5 V to 15 V on the device side of the bridge rectifier. Transients as high as 100 V will not cause system failure, although the encoder will not operate correctly under these conditions. Correct operation will resume immediately after the transient is removed. The output voltage of the synthesized sine wave is almost directly proportional to the supply voltage ($V_{DD} - V_{SS}$) and will increase with the increase of supply voltage between terminals 8 and 16, after which the output voltage remains constant.

DIAL PULSE INDICATOR

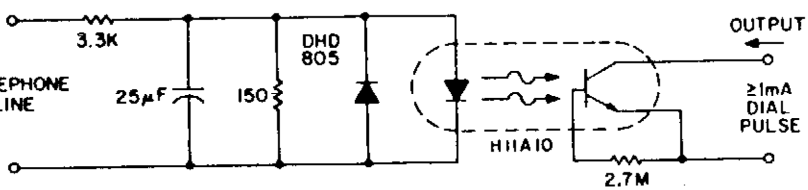
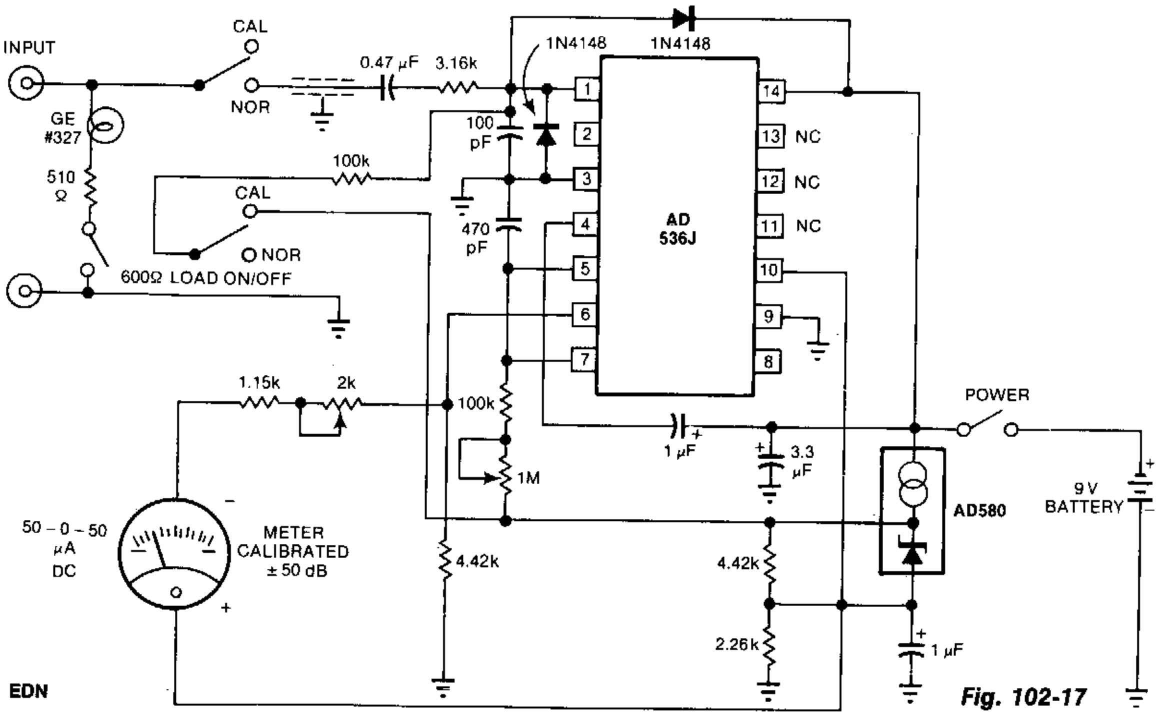


Fig. 102-16

This indicator senses the switching on and off of the 48-Vdc line voltage and transmits the pulses to logic circuitry. An H11A10 threshold coupler, with capacitor filtering, gives a simple circuit which can provide dial pulse indication, and yet reject high levels of induced 60-Hz noise. The DHD805 provides reverse bias protection for the LED during transient over-voltage situations. The capacitive filtering removes less than 10 ms of the leading edge of a 40-V dial pulse, while providing rejection of up to 25-V rms at 60 Hz.

TELEPHONE SOUND LEVEL METER MONITOR



EDN

Fig. 102-17

The telephone-line decibel meter and line-voltage sensor shown lets you accurately monitor and adjust telephone sound levels. The 600-Ω resistor properly terminates the line. Power drain from the 9-V battery is 2 mA, and the meter provides ±30 dB range.

REMOTE TELEPHONE RINGER

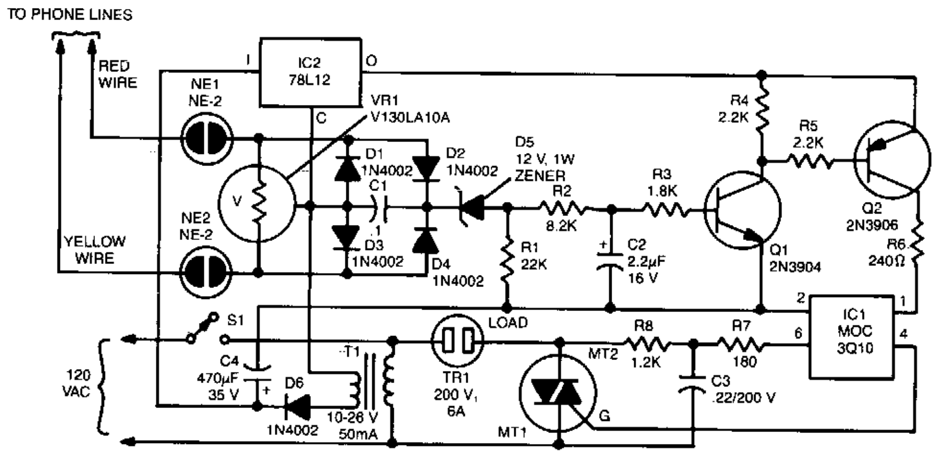
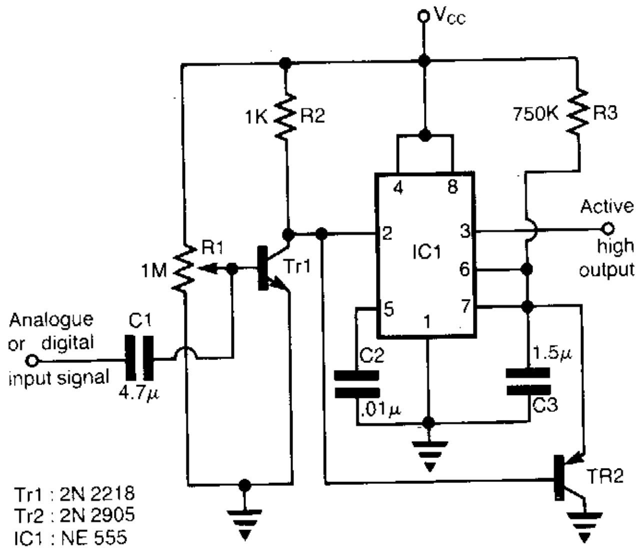


Fig. 102-18

The two neon bulbs will light when more than 100 V is across the ringing circuit. The bulbs provide line isolation between the unit and the telephone line. Finally, they act as a voltage divider for the bridge rectifier made up of D1 through D4. That voltage divider creates a positive voltage that is then applied through D5, is filtered by R2, R3, and C2, and causes Q1 and Q2 to conduct. When that happens, triac TR1 is fired through the optical coupler IC1; this turns on the triac, which applies 110 Vac to the load.

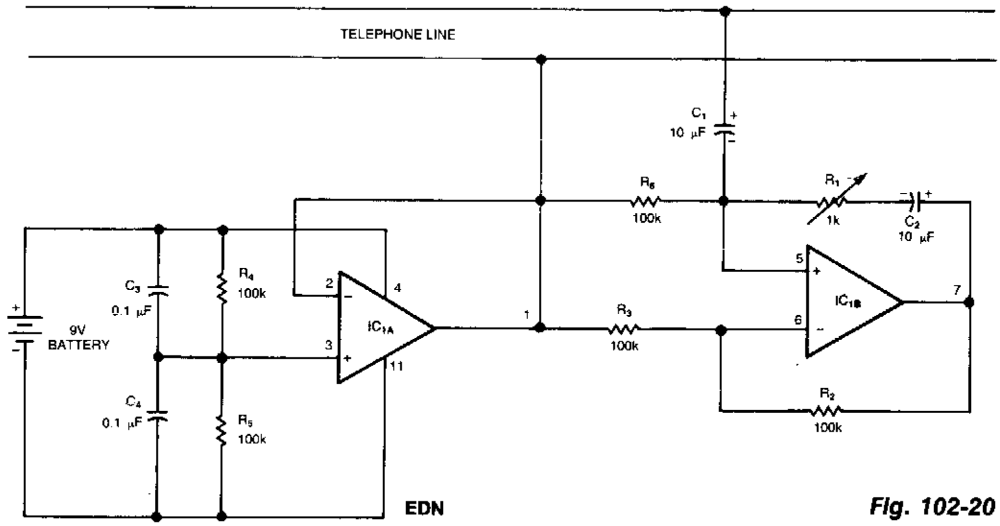
TELEPHONE SPEECH ACTIVITY DETECTOR



ELECTRONIC ENGINEERING *Fig. 102-19*

This circuit can be used in telephone lines for speech activity detection purposes. This detection is very useful in the case of half-duplex conversation between two stations—in the case of simultaneous transmission of voice and data over the same pair of cables by the method of interspersion data on voice traffic, and also in echo suppressor devices. The circuit consists of a class-A amplifier to amplify the weak analog signals (25 – 400 mV). The IC1 which follows, is connected as a retriggerable monostable multivibrator with the TR2 discharging the timing capacitor C3, if the pulse train reaches the trigger input 2 of IC1 with period less than the time: $T_{HIGH} = 1.1$ R3C3. The output 3 of IC1 is active on when an analog or digital signal is presented at the output, and it drops to a low level, T_{HIGH}, seconds after the input signal has ceased to exist.

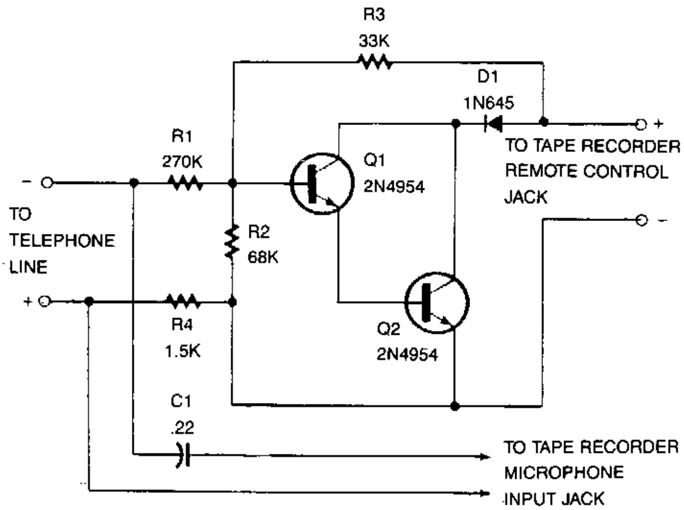
DUPLEX LINE AMPLIFIER



EDN *Fig. 102-20*

This circuit is a bidirectional amplifier that can amplify both signals of a duplex telephone conversation. It uses the principle of negative resistance. Obviously, such an amplifier could easily be unstable; however, you can adjust R1 for maximum amplification and the circuit will remain stable. The LM324 op amps can be replaced with op amps that would distort less, such as the LM1558, LF412, LF353, or LF442.

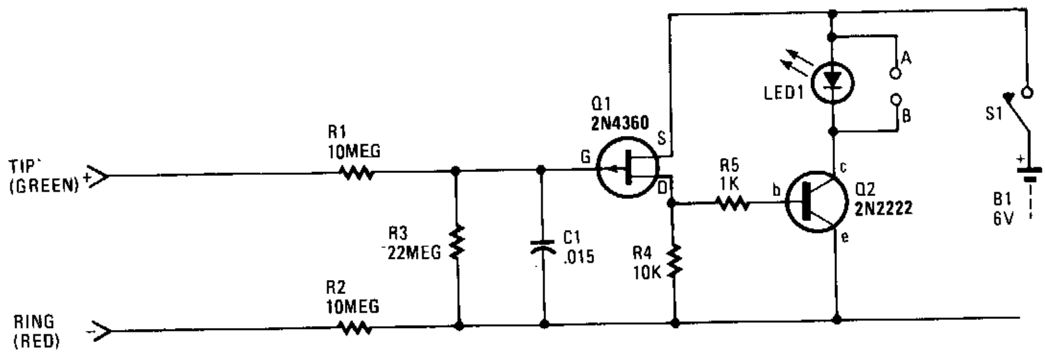
PHONE RECORDER



HANDS-ON ELECTRONICS *Fig. 102-21*

This recorder can be connected to the telephone lines just about any place, and no external power source is needed. The tape recorder's switch terminals are applied to a pair of transistors, connected as Darlingtons, that are used to turn the recorder on and off. When the telephone is off-hook there's usually about 50 Vdc across the phone that's divided over R1, R2, and R4, so that Q1's base is negative enough to keep the recorder off. Pick up the receiver, and the voltage drops to 5 V. That leaves not quite-enough voltage on Q1's base to keep that transistor at cutoff, so the recorder begins. Remember to keep your recorder's switch in the on position, and depending on how many people use the telephone, remember to rewind or change tapes occasionally!

LINE-ACTIVATED SOLID-STATE SWITCH



HANDS-ON ELECTRONICS

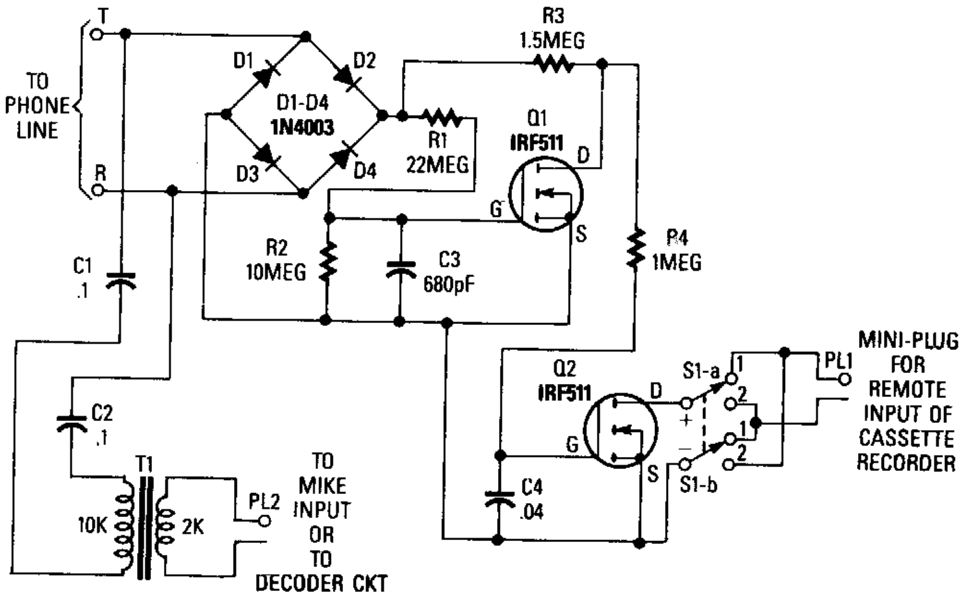
Fig. 102-22

Each and every time a phone on the same line or calling number is taken off-hook, the circuit will be activated to control an external electronic circuit. If several extension telephones are used on one phone line, the circuit can be useful as a *busy* indicator. LED1 contains a special flashing red LED that makes an excellent indicator for a *busy* circuit condition.

The solid-state switch can be used for several other phone-activated applications, such as automatically turning on a cassette recorder, starting a phone-use timer or counter, etc. A small relay can be connected at points A and B, in place of LED1, to control external circuits. A 117-Vac-to-6-Vdc plug-in power supply can be substituted for the battery to keep the operating cost at a minimum.

The 48-Vdc, on-hook, phone-line voltage keeps Q1 in the cut-off condition, allowing no current to flow through resistor R4, hence Q2 remains off. Resistors R1 and R2 keep the solid-state switch circuit from causing any problems with the telephone's central-office equipment. When a phone is taken off-hook, the line voltage (tip to ring) drops to 10 V or less, which forces Q1 to turn on; this, in turn, causes Q2 to trigger LED1, or a relay which might be used in lieu of LED1.

CASSETTE INTERFACE



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 102-23*

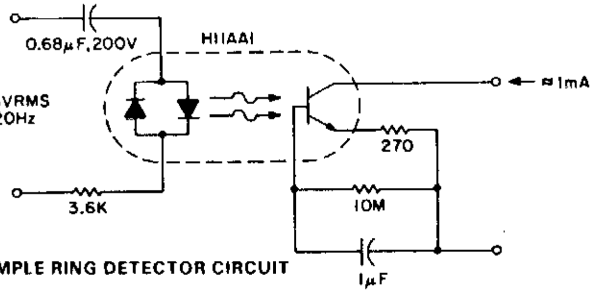
Q1 and Q2 are used to form the basis of an interface circuit for attaching a cassette recorder to the phone line. The circuit does not require a power supply because operating power is drawn from the telephone line itself. The incoming signal is fed across a bridge-rectifier circuit, consisting of diodes D1 through D4.

When the phone is on hook, the voltage at the output of the bridge at the R1/R3 junction is near 48 V. That voltage is fed across a voltage divider consisting of R1 and R2. The voltage at the junction formed by R1 and R2 is fed to the gate of Q1, turning it on. That pulls the drain of Q1 low. Since the gate of Q2 is connected to the drain of Q1, the bias applied to the gate of Q2 is low, holding it in the OFF state.

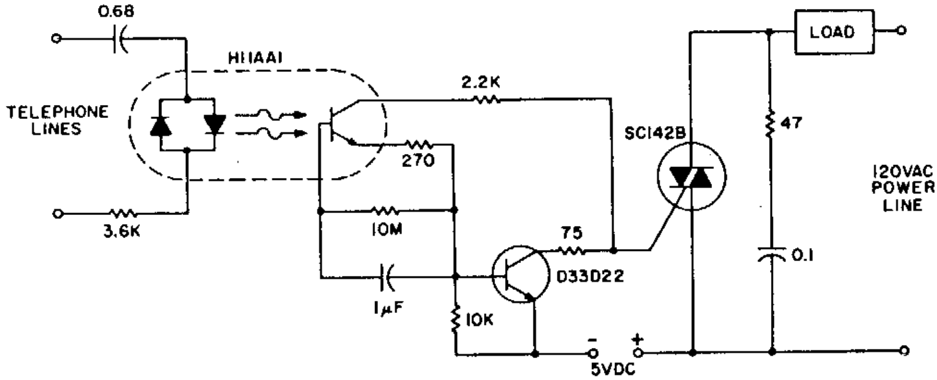
When the answering machine responds to a call or a phone is taken off hook, the voltage across the phone lines drops below 10 V, causing Q1 to turn off. At that point, the voltage at Q1's drain rises, turning Q2 on. The remote input of the cassette is connected to Q2's drain and source through S1, and a miniature plug is connected to the remote input jack.

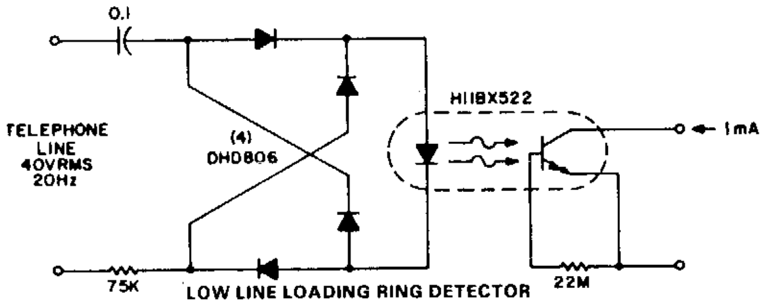
Switch S1 must be in a position so that the positive lead of the recorder's remote input connects, through switch position 1, to Q2's drain and the negative input to Q2's source. Switch S1 provides a convenient way to reverse the circuit's trigger output without having to unsolder and resolder leads. The phone's audio is coupled through C1, C2, and T1 to the microphone input of the cassette recorder.

RING DETECTOR

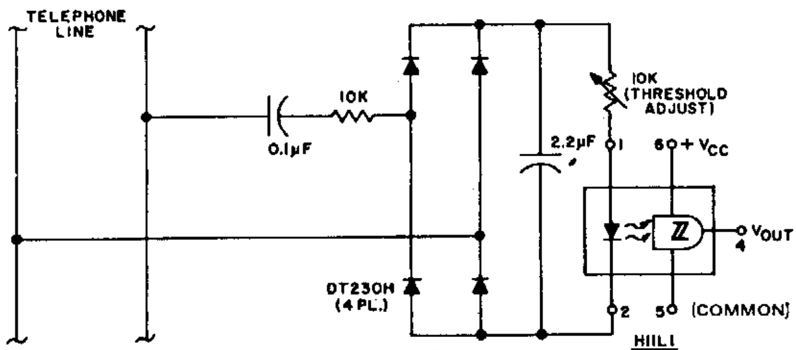


SIMPLE RING DETECTOR CIRCUIT





LOW LINE LOADING RING DETECTOR



RING DETECTOR USING H11L1

GE

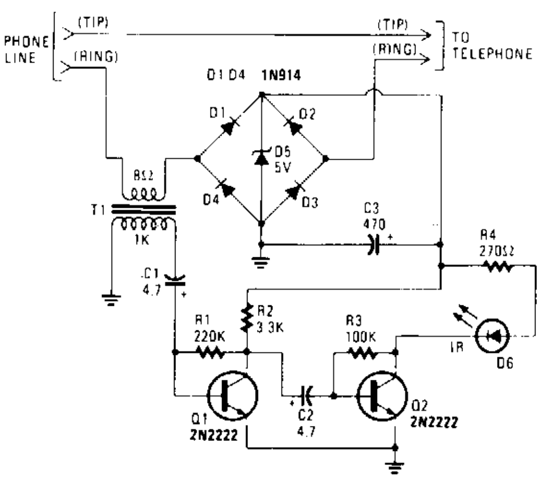
Fig. 102-24

RING DETECTOR (*Cont.*)

This circuit detects the 20 Hz, ≈ 86-V rms ring signal on telephone lines and initiates action in an electrically isolated circuit. Typical applications would include automatic answering equipment, and interconnect/interface and key systems. The circuits illustrated are *bare bones* circuits designed to illustrate concepts. They might not eliminate the ac/dc ring differentiation, 60-Hz noise rejection, dial tap rejection, and other effects that must be considered in field application. The first ring detector is the simplest and provides about 1-mA signal for a 7-mA line loading for 1/10 sec after the start of the ring signal. The time delay capacitor provides a degree of dial tap and click suppression, as well as filtering out the zero crossing of the 20-Hz wave. This circuit provides the basis for a simple example, a ring extender that operates lamps and buzzers from the 120-V, 60-Hz power line, while maintaining positive isolation between the telephone line and the power line. Use of the isolated tab triac simplifies heat sinking by removing the constraint of isolating the triac heatsink from the chassis. Lower line current loading is required in many ring detector applications. This can be provided by using the H11BX522 photo-Darlington optocoupler, which is specified to provide a 1-mA output from a 0.5-mA input through the −25°C to +50°C temperature range.

The next circuit allows ring detection down to a 40-V rms ring signal while providing 60-Hz rejection to about 20-V rms. Zero-crossing filtering can be accomplished either at the input bridge rectifier or at the output. Dependable ring detection demands that the circuit responds only to ring signals, rejecting spurious noise of similar amplitude, such as dialing transients. The configuration shown relies on the fact that ring signals are composed of continuous frequency bursts, whereas dialing transients are much lower in repetition rate. The dc bridge-filter combination at the H11L input has a time constant; it cannot react to widely spaced dialing transients, but will detect the presence of relatively long duration bursts, causing the H11L to activate the downstream interconnect circuits at a precisely defined threshold.

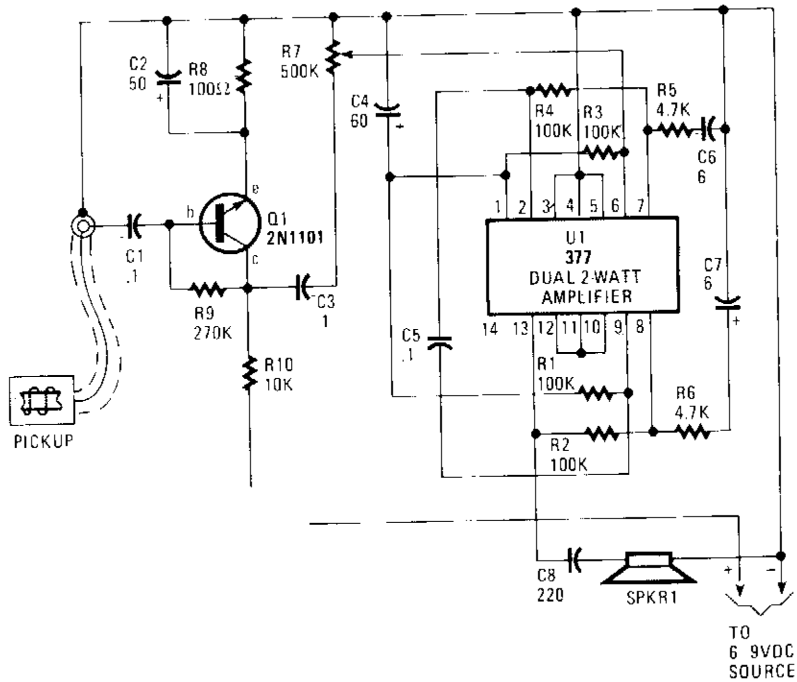
WIRELESS TELEPHONE EAVESDROPPER



HANDS-ON ELECTRONICS *Fig. 102-25*

The IR transmitter connects to a telephone circuit, and transmits both sides of all telephone conversations to any line-of-sight location, within 40 feet. No power is taken from the central office, as long as all phones remain on-hook. The current flows through the phone and back to the central office, thereby keying their equipment. We tap into the telephone line by connecting the IR transmitter circuit in series with either the tip or ring. When the telephone is off-hook, current will flow through the diode bridge polarity protector and supply the power for the IR transmitter. The phone's audio information is taken off the line by transformer T1. The 1000-Ω winding of the transformer connects to a two-stage transistor audio amplifier/modulator. A 2000-Ω potentiometer could be added to the input of the two-stage amplifier to control the modulation level, and another potentiometer could be added in place of R3 to adjust the IR's idle current.

TELEPHONE AMPLIFIER

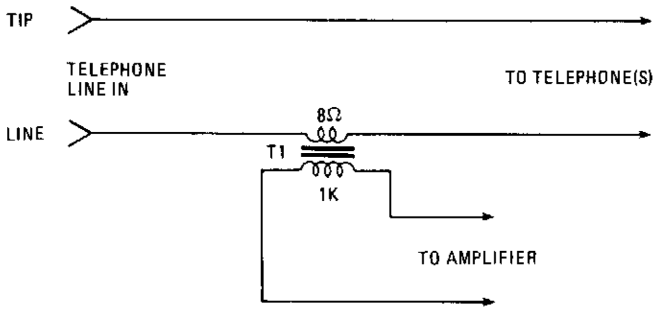


TAB BOOKS

Fig. 102-26

Audio from the telephone is inductively coupled to the base of Q1, which is used as a preamp. The preamp provides a gain of about 75 dB, to boost the input signal from about 4 mV to about 300 mV pk-pk. If you use a higher gain transistor, increase the value of R9 to produce a *Q* point, measured from minus to the collector of Q1, of one-half the supply voltage. The Q1 output signal is coupled through C3 to R7, which serves as a volume or drive-level control, to U1, a dual, 2 w amplifier connected in cascade. Pins 1 through 7 serve as a driver for the final amplifier, pins 8 through 13. Compensation and balance is accomplished by components R1 through R6 and C4, C6, and C7. Pins 3 through 5, and 10 through 12 should be tied to the negative supply rail.

TELEPHONE TAP



HANDS-ON ELECTRONICS ***Fig. 102-27***

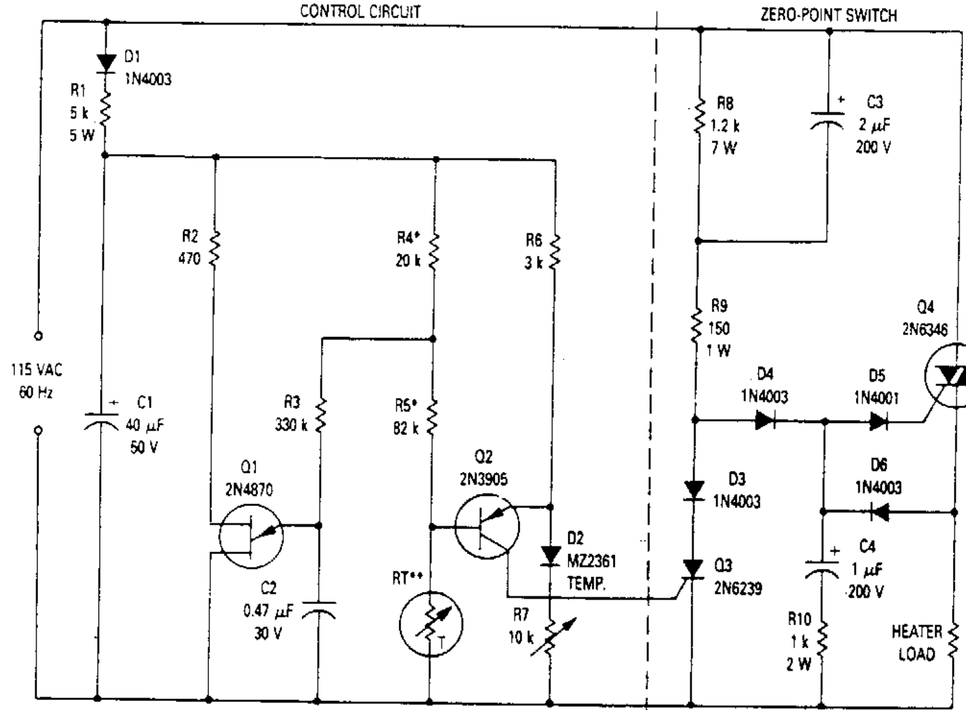
Amplify or record a telephone call with the simple circuit shown. The 8-Ω secondary winding of a miniature transistor output transformer is connected in series with either of the telephone lines. The 1000-Ω primary winding can feed either a cassette recorder or an audio amplifier.

103

Temperature Controls

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

ZERO-POINT SWITCHING TEMPERATURE CONTROL

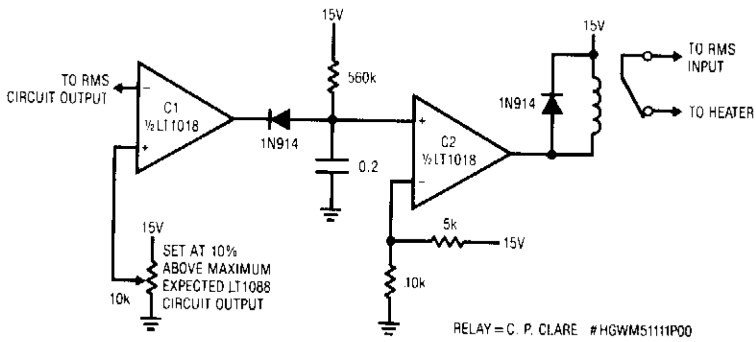


*LOW TEMP. COEFFICIENT
**FENWELL GR51J1 100 k THERMISTOR

Fig. 103-1

This modulated triac zero-point switching circuit controls heater loads operating from 115 Vac. Circuit operation is best described by splitting the circuit into two parts. The circuit at right is the zero-point switch; to the left is the proportional control for the zero-point switch.

SERVO-SENSED HEATER PROTECTOR

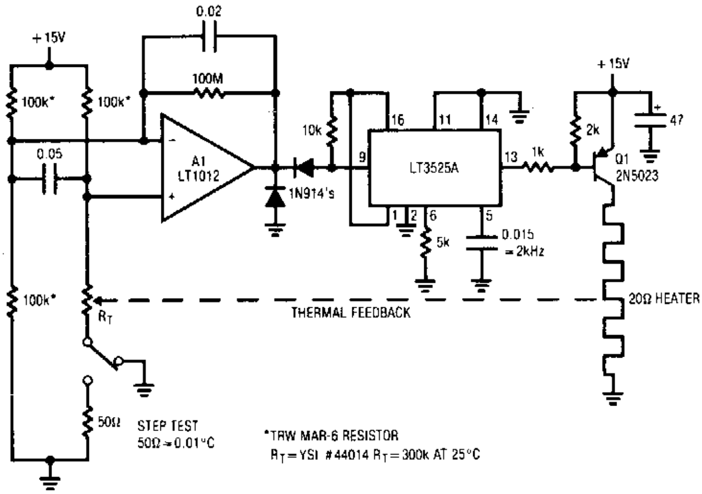


LINEAR TECHNOLOGY TECH.

Fig. 103-2

This circuit responds quickly enough to prevent damage from most overloads. C1's input is connected to the output of the LT1088 servo circuit. If the LT1088 circuit's output exceeds the threshold at C1's other input, C1 trips, discharging the 2-μF capacitor. This causes C2's output to become low, energize the relay, and break the heater circuit. The 560-KΩ resistor provides a long recharge for the capacitor, preventing chattering action. This arrangement's speed of response is limited by the rms circuit's slew rate, about 0.2 V/ms. For reasonable overloads, the LT1088's temperature increases about 1°C/ms. A 10-V LT1088 output step takes 50 ms, causing a temperature rise of about 50°C.

TEMPERATURE CONTROLLER



LINEAR TECHNOLOGY CORP. ***Fig. 103-3***

When power is applied, the thermistor, a negative tc device, is at a high value. A1 saturates positive. This forces the LT3525A switching regulator's output low, biasing Q1. As the heater warms, the thermistor's value decreases. When its inputs finally balance, A1 comes out of saturation and the LT3525A pulse-width modulates the heater via Q1, completing a feedback path. A1 provides gain and the LT3525A is highly efficient. The 2-kHz, pulse-width modulated heater power is much faster than the thermal loop's response, and the oven sees an even, continuous heat flow.

PROPORTIONAL TEMPERATURE CONTROLLER

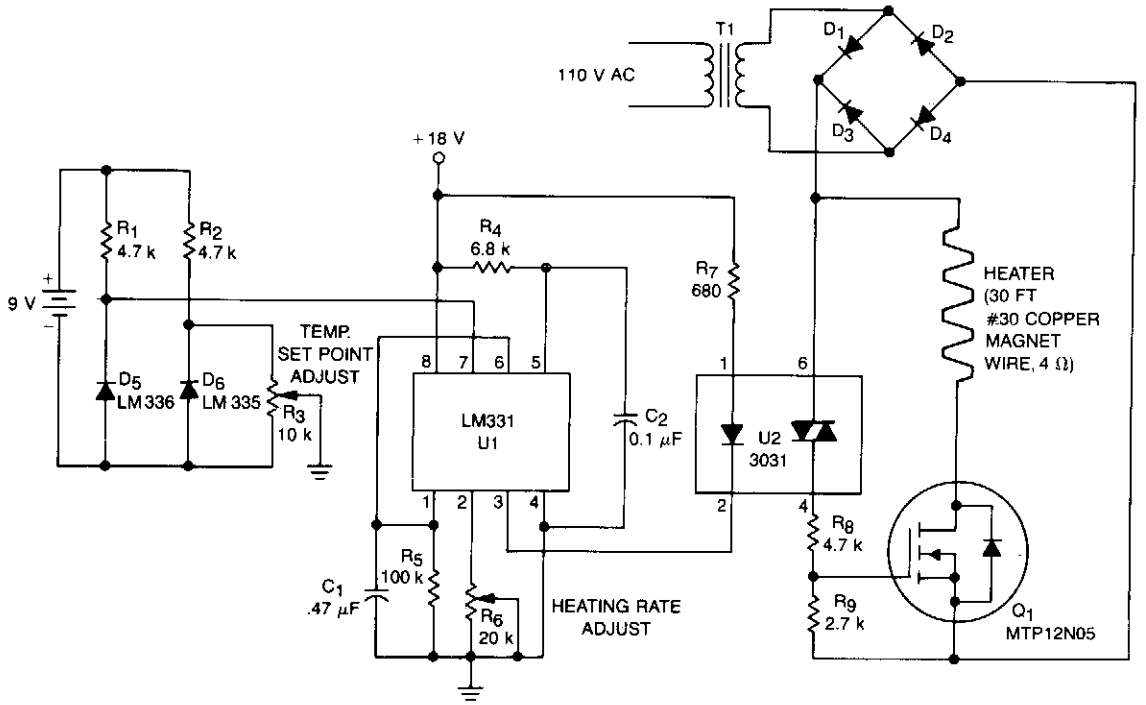
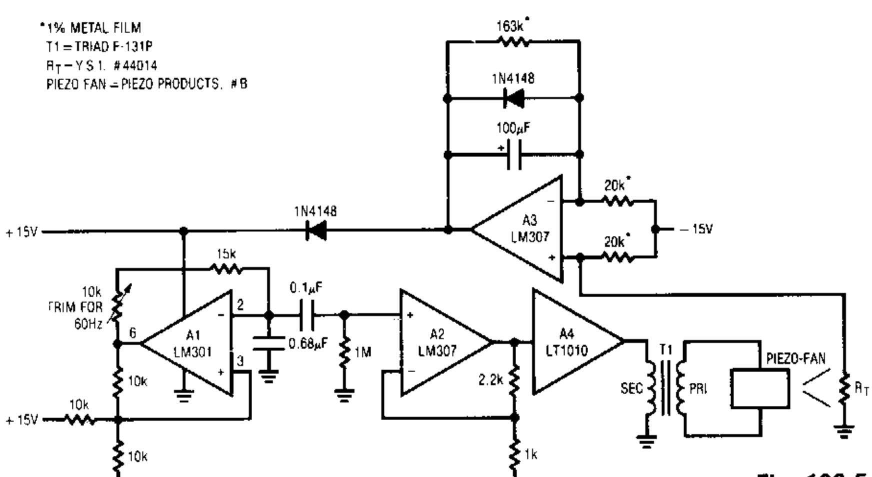


Fig. 103-4

This temperature controller operates as a *pulse snatching* device, which allows it to run at its own speed and turn on at the zero crossing of the line frequency. Zero crossing turn-on reduces the generation of line noise transients. TMOS Power FET, Q1, is used to turn on a heater.

Temperature sensor D6 provides a dc voltage proportional to temperature that is applied to voltage-to-frequency converter U1. Output from U1 is a pulse train proportional to temperature offset that is applied to the input of triac optoisolator U2. The anode supply for the triac is a 28 V pk-pk, full-wave rectified sine wave. The optoisolator ORs the pulse train from U1 with the zero crossing of U2's anode supply, supplying a gate turn on signal for Q1. Therefore, TMOS power FET Q1 can only turn the heater on at the zero crossing of the applied sine wave. The maximum temperature, limited by the sensor and the insulation of the wire, is 130°C for the components shown.

PIEZOELECTRIC FAN-BASED TEMPERATURE CONTROLLER



LINEAR TECHNOLOGY CORP.

Fig. 103-5

The fan employed is one of the new electrostatic type which is very reliable, because it contains no wearing parts. These devices require high-voltage drive. When power is applied, the thermistor, located in the fan's exhaust stream, is at a high value. This value unbalances the A3 amplifier driven bridge. A1 receives no power and the fan does not run. As the instrument enclosure warms, the thermistor value decreases until A3 begins to oscillate. A2 provides isolation and gain, and A4 drives the transformer to generate high voltage for the fan. In this fashion, the loop acts to maintain a stable instrument temperature by controlling the fan's exhaust rate. The 100-μF time constant across the error amplifier pins is typical of such configurations. Fast time constants will produce audibly annoying *hunting* in the servo. Optimal values for this time constant and gain depend upon the thermal and airflow characteristics of the enclosure being controlled.

ELECTRONIC HEAT SNIFFER

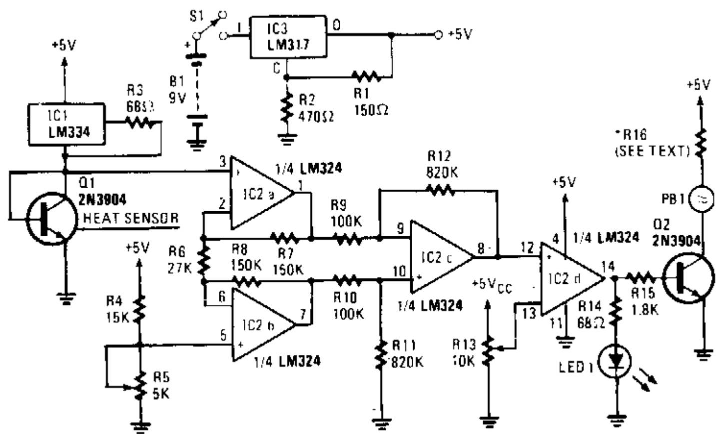


Fig. 103-6

Sensing element Q1 is a 2N3904 general-purpose npn transistor, although any general-purpose npn unit in a TO-92 style case will do. IC1, an LM334, supplies Q1 with a constant current that is independent of temperature. An LM324 quad op amp, IC2, forms a high input-impedance differential amplifier (IC2a, IC2b, and IC2c) with a gain of about 99. IC2d is used as a voltage comparator. When Q1 senses a rise or fall in temperature, the base-to-emitter voltage decreases. That decrease in voltage causes the input to IC2a at pin 3 to deviate from the reference voltage that's fed to IC2b at pin 5, which is set by potentiometers R5. The difference between the input and the reference is amplified by IC2c. That amplified voltage is fed to IC2d where it is compared to a control voltage set by potentiometer R13. The setting of R13 determines the threshold and is set at a point that's equal to the ambient temperature. The output of IC2d at pin 14 is fed to the base of transistor Q2. When the output of IC2d is high, LED1 lights and Q2 turns on. With Q2 turned on, a ground path through the transistor is provided for buzzer PB1.

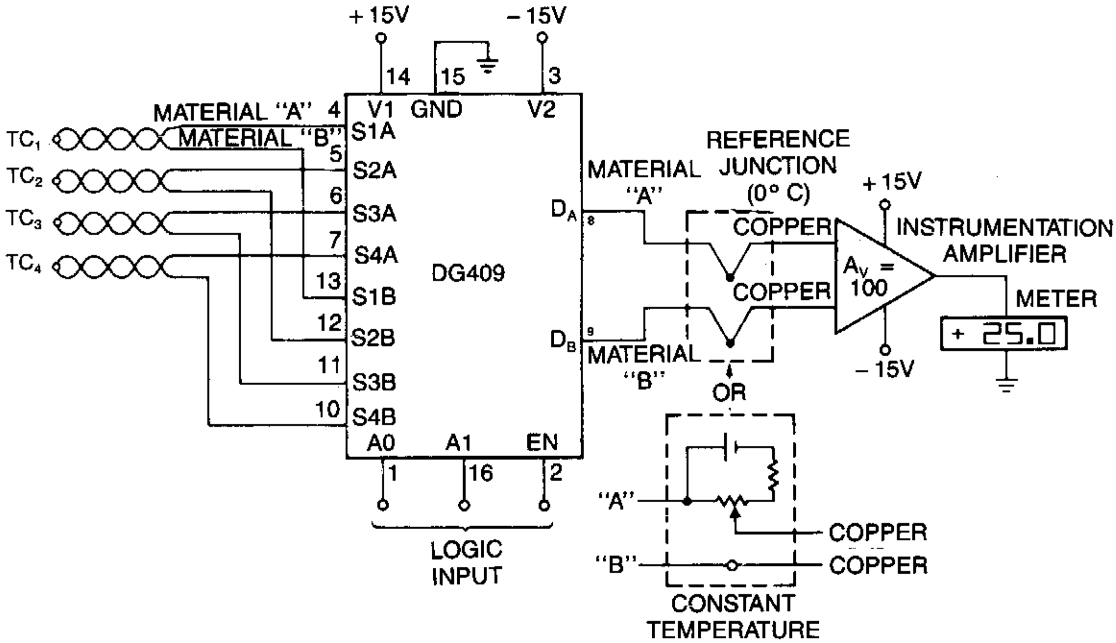
The circuit can be built on perforated construction board using point-to-point wiring. All components, except Q1, are mounted on the board. Transistor Q1 is mounted at the tip of the heat-sensing probe.

104

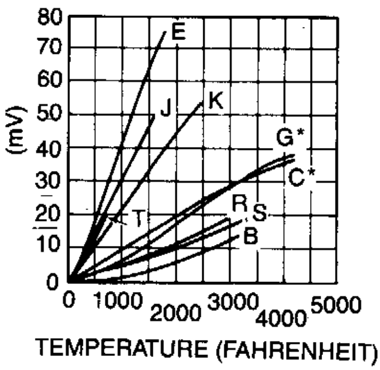
Temperature Sensors

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

THERMOCOUPLE MULTIPLEX SYSTEM



Output Voltage vs. Temperature of Several Common Thermocouples



ANSI SYMBOL	
T	Copper vs Constantan
E	Chromel vs Constantan
J	Iron vs Constantan
K	Chromel vs Alumel
G*	Tungsten vs Tungsten 26% Rhenium
C*	Tungsten 5% Rhenium vs Tungsten 26% Rhenium
R	Platinum vs Platinum 13% Rhodium
S	Platinum vs Platinum 10% Rhodium
B	Platinum 6% Rhodium vs Platinum 30% Rhodium

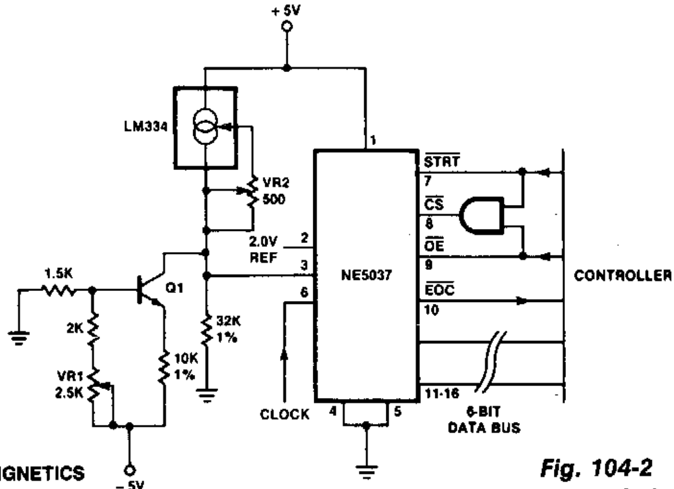
*Not ANSI Symbol

SILICONIX

Fig. 104-1

To decouple the sensors from the meter amplifier, either a reference junction at 0°C or a bucking voltage set at room temperature may be used. The latter method is simpler, but is sensitive to changes in ambient temperature. The table above shows the output voltage vs temperature of several common types of thermocouples.

O – 63 °C TEMPERATURE SENSOR

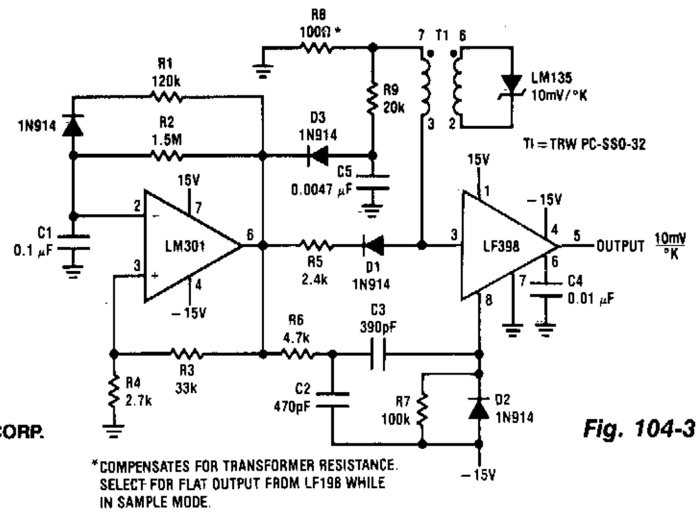


SIGNETICS

Fig. 104-2

The temperature sensor provides an input to pin 3 of the NE5037 of 32 mV/°C. This 32 mV is the value of one LSB for the NE5037. The LM334 is a three-terminal temperature sensor and provides a current of 1 μA for each degree Kelvin. The 32-KΩ resistor provides the 32 mV for each microamp through it, while the transistor bleeds off 273 μA of the temperature sensor (LM334) current. This bleeding lowers the reading by 273 K, thus converting from Kelvin to Celsius. To read temperature, conversion is started by sending a momentary low signal to pin 7 of the NE5037. When pin 10 of the NE5037 becomes low, conversion is complete and a low is applied to pin 9 of the NE5037 to read data on pins 11 and through 16. Note that this temperature data is in straight binary format. The controller can be a microprocessor in a temperature control application, or discrete circuitry in a simple temperature reporting application.

ISOLATED TEMPERATURE SENSOR



LINEAR TECHNOLOGY CORP.

Fig. 104-3

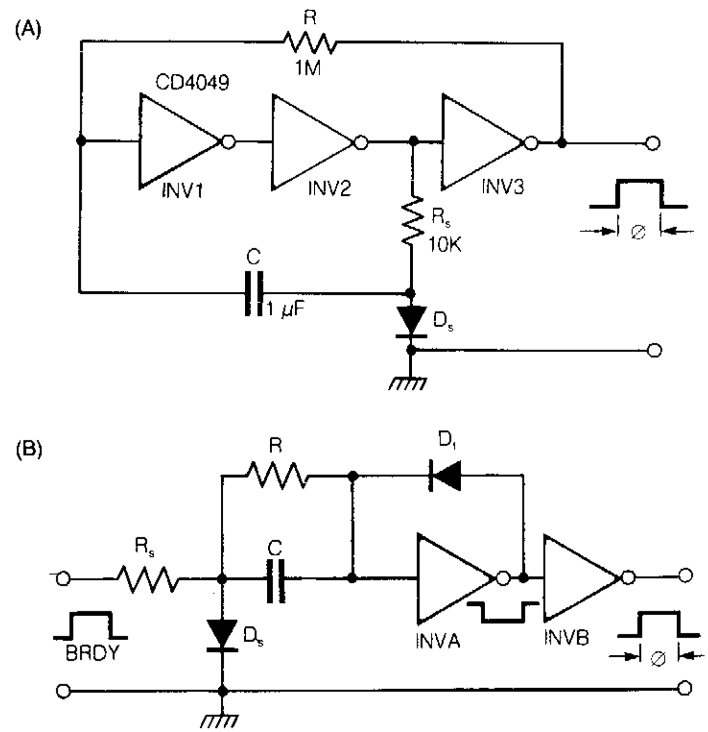
105

Temperature-to-Time Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Two Simple Temperature-to-Time Converters

TWO SIMPLE TEMPERATURE-TO-TIME CONVERTERS



ELECTRONIC ENGINEERING ***Fig. 105-1***

Both of these converters use CMOS inverters. Figure 105-1A shows a free-running circuit having both the pulse duration and pulse pause dependent on temperature of the diode D_S. It can be used where a synchronization between the converter and something else is not required.

Figure 105-1B shows a one shot circuit that produces a pulse with its duration dependent of temperature of diode D_S. The additional diode D_f should have inverse current low enough to not influence the discharging process in the network rc when the INVA output is low. A silicon component or a GaAsP LED can be used.

The converter is intended for a digital system producing a RADY pulse which disappears after the conversion process is ended. The pulse duration is approximately:

$$= 2RC \frac{V_D}{V_{DD}}$$

where V_D is the sensor diode forward voltage and V_{DD} is the supply voltage of the CMOS chip.

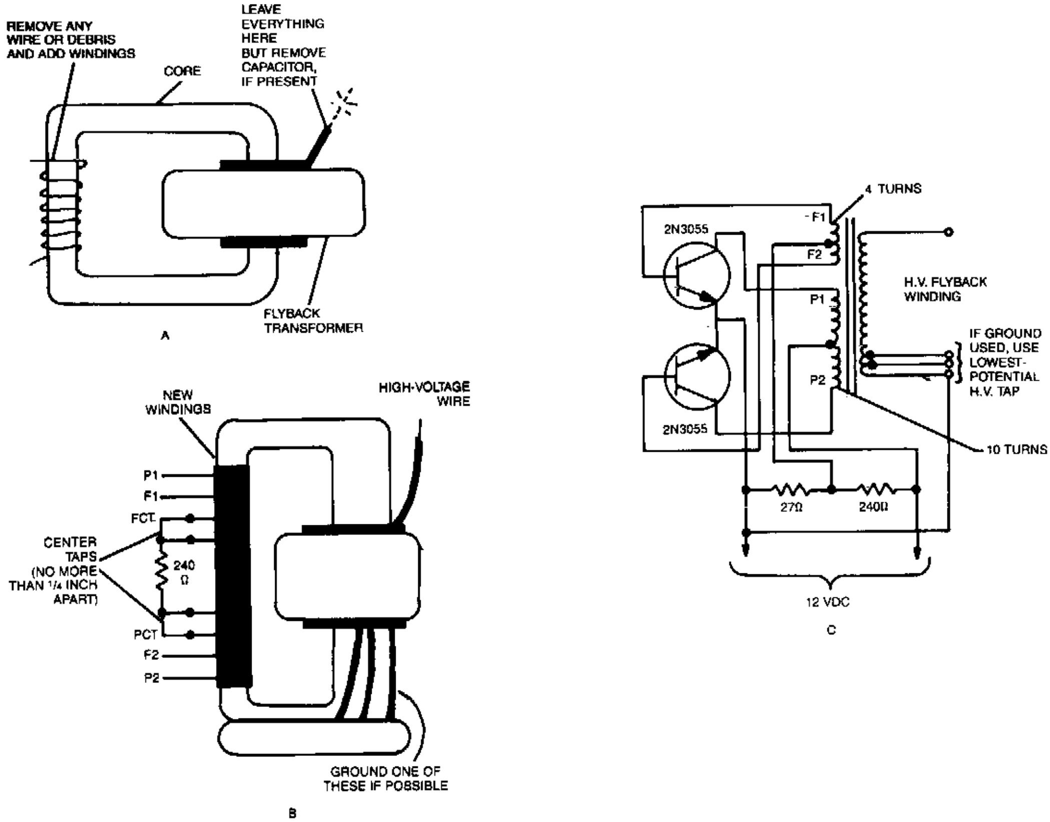
Resistance R must be much higher than R_S. A 0.1-μF capacitor can be applied in parallel with D_S, if necessary, to repulse stray pickup and noise in a long cable. The circuits described can be used with a temperature sensitive resistor instead of the diode D_S.

106

Tesla Coils

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SIMPLE TESLA COIL



GERNSBACK PUBLICATIONS INC.

Fig. 106-1

The Tesla coil described here can generate 25,000 V. So, even though the output current is low, **be very careful!** The main component is a flyback transformer from a discarded TV.

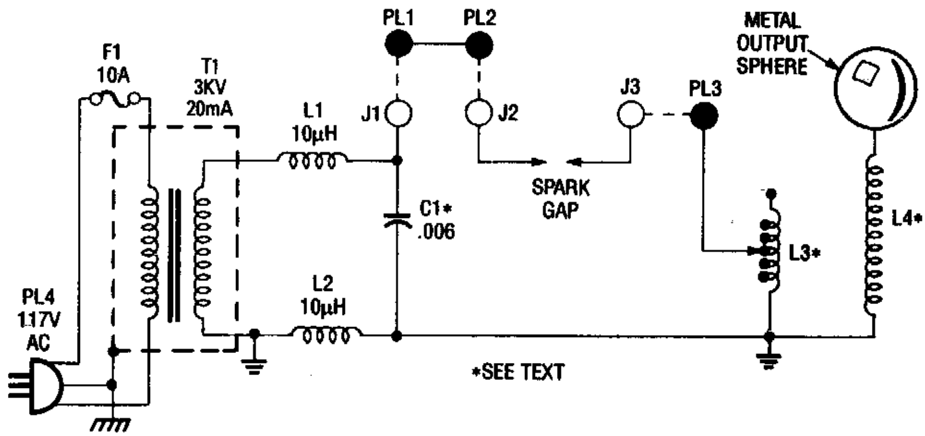
A new primary winding is needed. Begin by winding 5 turns of #18 wire on the core. Then, twist a loop in the wire, and finish by winding five more turns. Wrap with electrical tape, but leave the loop exposed.

A four-turn winding must be wound over the ten-turn winding that you've just finished. That is done the same way. First wind two turns of # 18 wire, then make a loop, and finish by winding two more turns. Again, wrap the new winding with electrical tape, leaving the loop exposed.

When the windings are finished, the two loops shouldn't be more than 1/4-inch apart, but take care that they do not touch. Connect a 240-Ω resistor between the two loops. The modified transformer now should look like the one shown. Connect the transformer as shown. The 27-Ω resistor and two transistors should be mounted on a heatsink and must be insulated from it.

The output of the high-voltage winding should begin to oscillate as soon as the circuit is connected to a 12-Vdc power supply. If it does not, reverse the connections to the base leads of the transistors. In normal operation, you should be able to draw 1-inch sparks from the high-voltage lead using an insulated screw-driver.

TESLA COIL



POPULAR ELECTRONICS

Fig. 106-2

Power is fed to transformer T1, a small neon-sign transformer, which steps the voltage up to about 3000 Vac. The stepped-up output of T1 is fed through L1 and L2 and across C1, causing the capacitor to charge until enough power is stored in the unit to produce an arc across the spark gap. The spark gap, which momentarily connects C1 and L3 in parallel, determines the amount of current transferred between C1 and L3.

The arcing across the spark gap sends a series of high-voltage pulses through L3, giving a sort of oscillated effect. The energy fed through L3 is transferred to L4 via the magnetic coupling between the two coils. Because of the turn ratio that exists between L3 and L4, an even higher voltage is produced across L4. Coil L4 steps up the voltage, which collects on the top-capacitance sphere. There, it causes an avalanche breakdown of the surrounding air, giving off a luminous discharge.

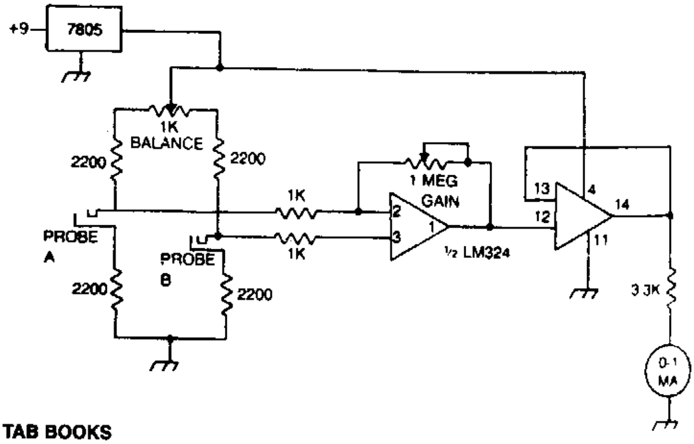
The rotary spark gap is a simple add-on circuit for the Tesla Coil, consisting of a variable dc power supply and a small, 5000-rpm, dc motor. The circuit allows you to vary the output of the Tesla coil by adjusting the rotating speed of the motor. A rotary gap is far more efficient than a stationary gap, because the stationary gap could cut-out and require readjustment.

107

Thermometer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DIFFERENTIAL THERMOMETER

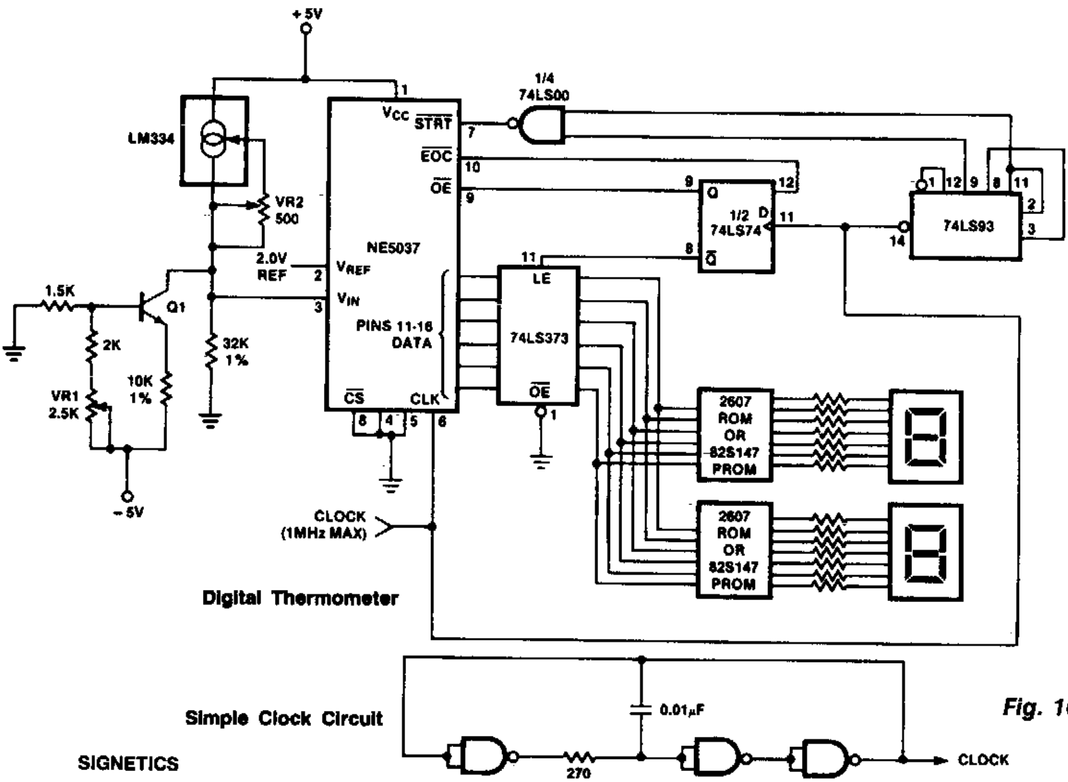


TAB BOOKS

Fig. 107-1

The differential thermometer uses two probes and shows the temperature difference between them, rather than the exact temperature. The thermometer uses a conventional meter as an indicator, and it covers a total range of 20° – 10°low to 10°high.

TEMPERATURE-REPORTING DIGITAL THERMOMETER

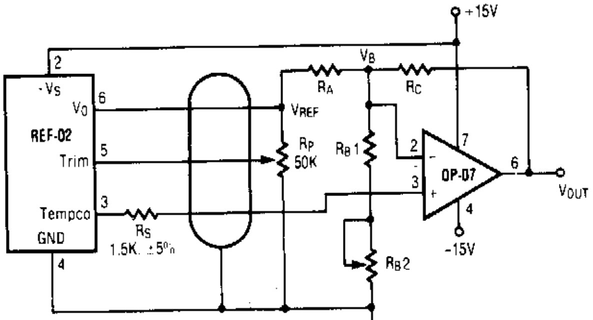


SIGNETICS

Fig. 107-2

The ROMs or PROMs must have the correct code for converting the data from the NE5037—used as address for the ROMs or PROMs—to the appropriate segment driver codes. The displayed amount could easily be converted to degrees Fahrenheit, °F, by the controller of (0 – 63°temperature sensor) or through the (P) ROMs. When doing this, a third (hundreds) digit (P)ROM and display will be needed for displaying temperatures above 99°F. An expensive clock can be made from NAND gates or inverters as shown.

ELECTRONIC THERMOMETER



*Up to 10 feet of shelded 4-conductor cable

$T_CV_{OUT} = (2.1 mV/°C)\left(1 + \frac{R_C}{R_A + R_B}\right)$

$V_O = \left(1 + \frac{R_C}{R_A + R_B}\right)V_{Tempco} - \left(\frac{R_C}{R_A}\right)(V_O)$

Resistor Values

TCV_{OUT} Slope(s)	10mV/°C	100mV/°C	10mV/°F
Temperature Range	−55°C to +125°C	−55°C to +125°C	−65°F to +257°F
Output Voltage Range	−0.55V to +1.25V	−5.5V to +12.5V	−0.67V to +2.57V
Zero Scale	0V at 0°C	0V at 0°C	0V at 0°F
R_A (±1% Resistor)	9.09KΩ	15KΩ	8.25KΩ
R_{B1} (±1% Resistor)	1.5KΩ	1.82KΩ	1.0KΩ
R_{B2} (Potentiometer)	200Ω	500Ω	200Ω
R_C (±1% Resistor)	5.11KΩ	84.5KΩ	7.5KΩ

Fig. 107-3

Reprinted with permission from Raytheon Co., Semiconductor Division.

This circuit uses the +5 V reference output and the op amp to level shift and amplify the 2.1 mV/°C Tempco output into a voltage signal dependent on the ambient temperature. Different scaling can be obtained by selecting appropriate resistors from the table giving output slopes calibrated in degrees Celsius or degrees Fahrenheit. To calibrate, first measure the voltage on the Tempco pin, V_{TEMPCO}, and the ambient room temperature, T_A in °C. Put those values into the following equation:

$$\frac{V_{TEMPCO}\ (\text{in mV})}{(S)\ (T_A + 273)}$$

Where S = Scale factor for your circuit selected from the table in mV. Then turn the circuit power off, short V_{OUT} at pin 6 of the REF-02 to ground, and while applying exactly 100.00 mV to the op amp output, adjust R_{B2} to that V_B = (X) (100 mV). Now remove the short and the 100-mV source, reapply circuit power and adjust R_P so that the op-amp output voltage equals (T_A) (S). The system is now exactly calibrated.

TEMPERATURE MEASURING ADD-ON FOR DMM DIGITAL VOLTMETER

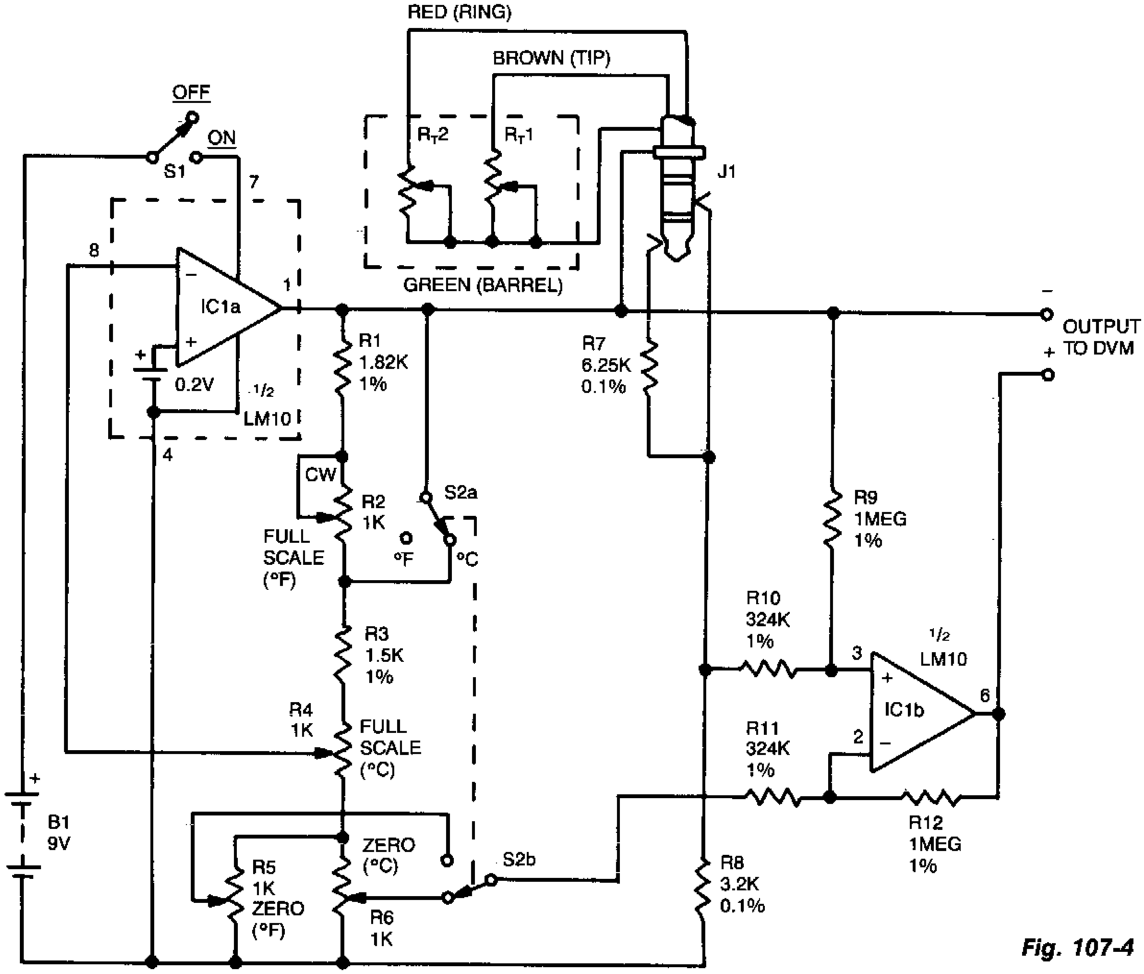


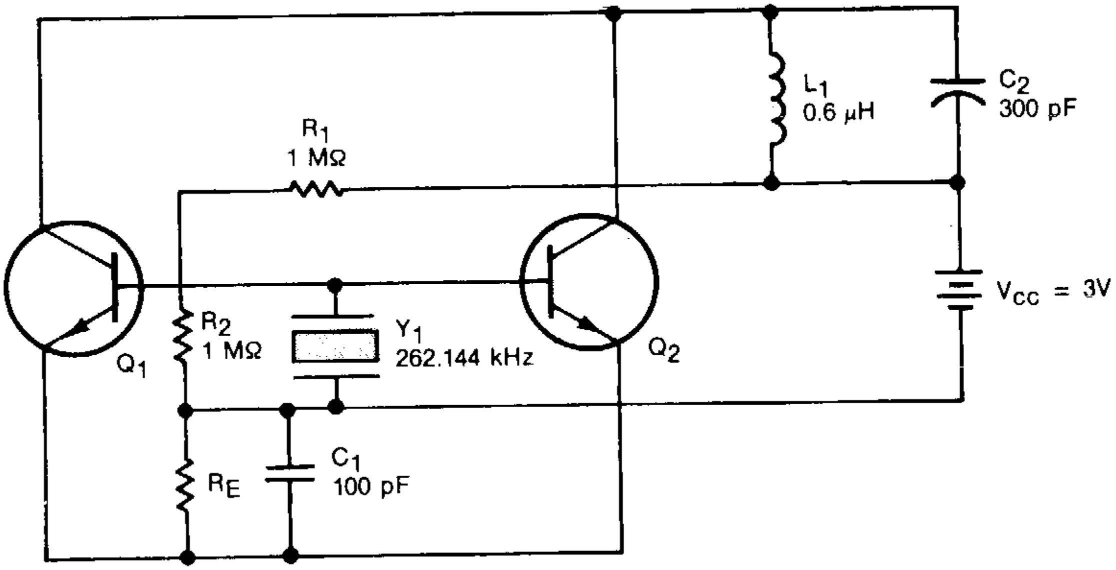
Fig. 107-4

The DVM-to-temperature adapter is built around a single IC, National's LM10. That micropower IC contains a stable 0.2 V reference, a reference amplifier and a general-purpose op amp. The circuit is designed for a linear temperature range of 0 to 100°C (32 to 212°F). The 0.2-V reference and reference amplifier provide a stable, fixed-excitation voltage to the Wheatstone bridge. The voltage is determined by a feedback network consisting of R1 through R6. Switch S2a configures the feedback to increase the voltage from 0.6 V on the Celsius range to 1.08 V on the Fahrenheit range. These differences compensate for the fact that one degree Fahrenheit produces a smaller resistance change than does one degree Celsius.

Resistors R1 through R16 also form the fixed leg of the Wheatstone bridge, nulling the bridge output at zero degrees. Since 0°C is different from 0°F, S2b is used to select the appropriate offset.

The LM10's op amp, along with R9 through R12, form a differential amplifier that boosts the bridge output to 10 mV per degree. Since a single supply is used, and since the output must be able to swing both positive and negative, the output is referenced to the bridge supply voltage, rather than to the common supply.

IMPLANTABLE INGESTIBLE ELECTRONIC THERMOMETER



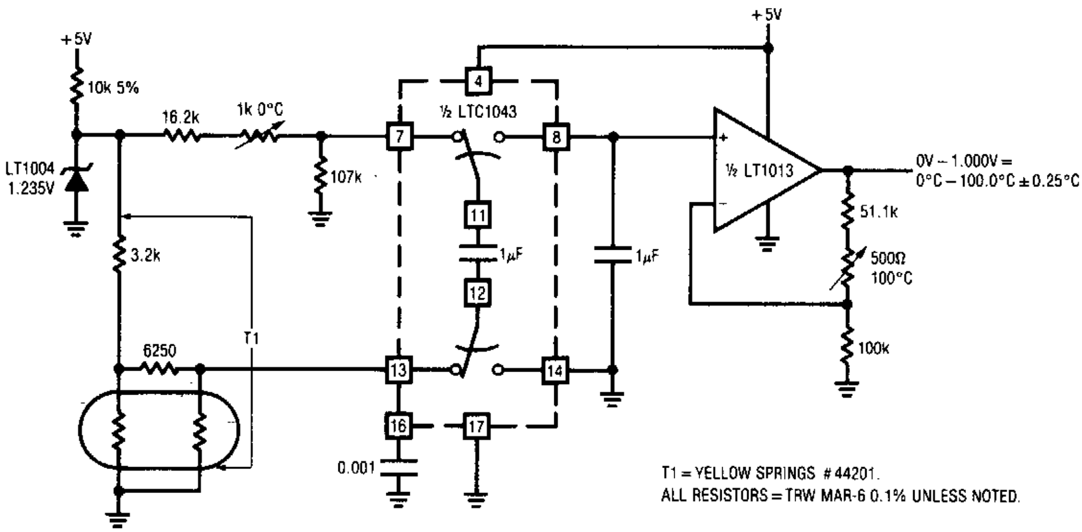
NOTES: 1. Q_1 and Q_2 are MAT-02 bipolar npn transistors.
2. R_E = 20 to 40 kΩ.

NASA

Fig. 107-5

This oscillator circuit includes a quartz crystal that has a nominal resonant frequency of 262,144 Hz and is cut in the orientation that gives a large linear coefficient of frequency variation with the temperature. In this type of circuit, the oscillation frequency is controlled primarily by the crystal—as long as the gain-bandwidth product is at least four times the frequency. In this case, the chosen component values yield a gain-bandwidth product of 1 MHz. Inductor L1 can be made very small: 100 to 200 turns with a diameter of 0.18 in. (4.8 mm) and a length of 0.5 in. (12.7 mm). Although the figure shows two transistors in parallel, one could be used to reduce power consumption or three could be used to boost the output. The general oscillator circuit can be used to measure temperatures from −10 to +140°C. A unit made for use in the human body from about 30 to 40°C operates at 262,144 ±50 Hz with a frequency stability of 0.1 Hz and a temperature coefficient of 9 Hz/°C.

SIMPLE LINEAR THERMOMETER

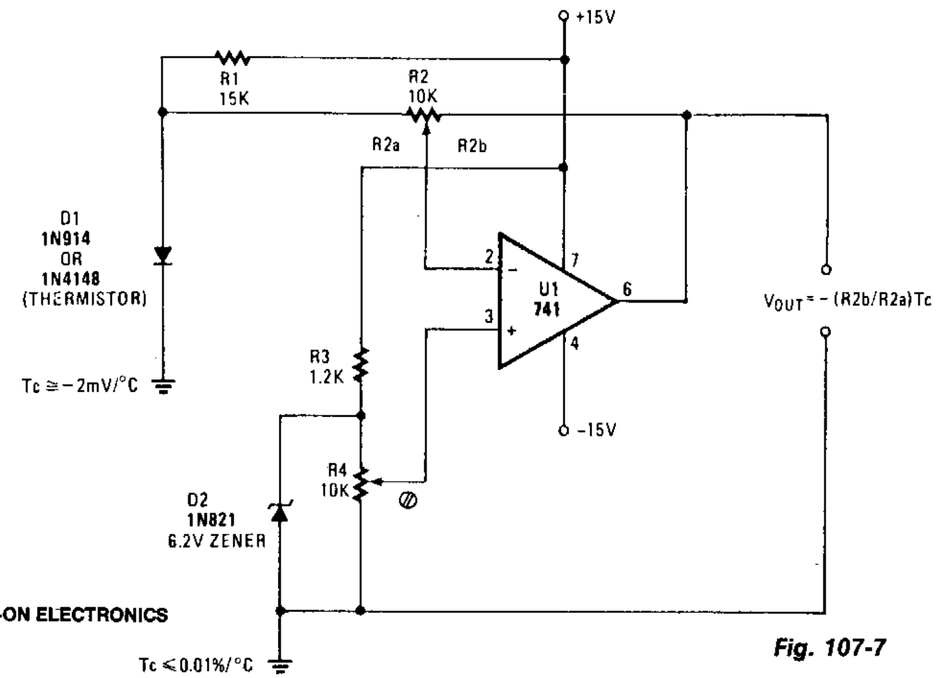


LINEAR TECHNOLOGY CORP.

Fig. 107-6

The thermistor network specified eliminates the need for a linearity trim—at the expense of accuracy and operational range.

THERMOMETER ADAPTER



HANDS-ON ELECTRONICS

Fig. 107-7

THERMOMETER ADAPTER (*Cont.*)

A simple op amp and silicon diode are the heart of the temperature-to-voltage converter that will permit you to use an ordinary voltmeter—either analog or digital—to measure temperature. User adjustments make it possible for a reading of either 10 mV or 100 mV to represent 1°F or C.

Temperature sensor D1 is a 1N4148 silicon diode. It has a temperature coefficient of −2 mV/°C. U1, a 741 op amp, is connected as a differential amplifier. A voltage divider consisting of R3 and Zener diode D2 provides a 6.2 V reference voltage. D2 is shunted by potentiometer R4, so that the offset can be adjusted to align the output voltage with either the Celsius or Fahrenheit scale, as desired.

Gain control R2 is adjusted so the output of the op amp is in the scale or voltage range of the meter being used. R4, the offset adjust control, is then adjusted so the output voltage represents either degrees F or C. The thermometer adapter can be calibrated by adjusting R4 while the probe sensor is at a known temperature.

108

Tilt Meter

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Digitizer

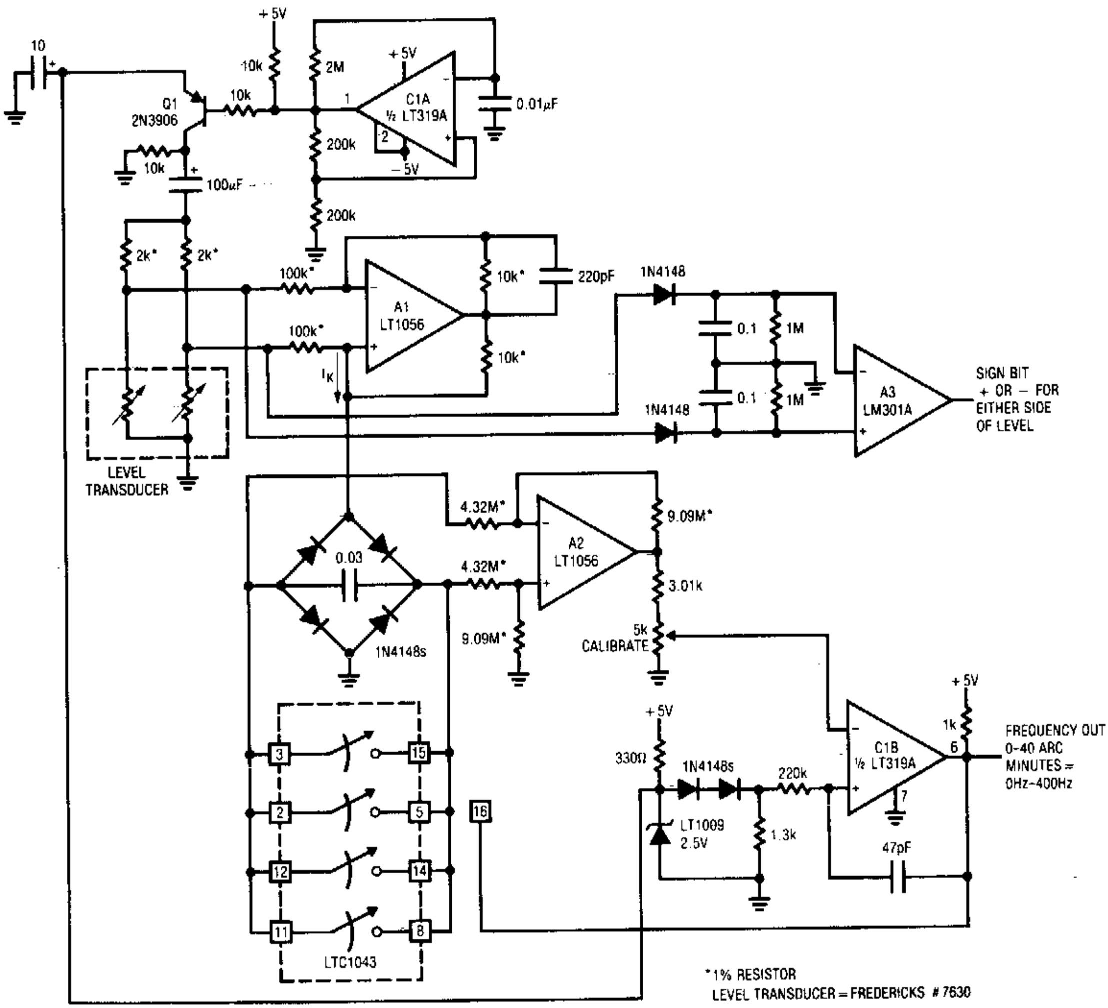


Fig. 108-1(a)

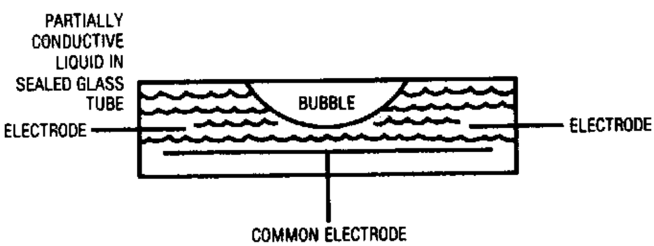


Fig. 108-1(b)

LINEAR TECHNOLOGY

Bubble Based Level Transducer

If the tube is level with respect to gravity, the bubble resides in the tube's center and the electrode resistances to common are identical. As the tube shifts away from level, the resistances increase and decrease proportionally. Transducers of this type must be excited with an ac waveform to avoid damage to the partially conductive liquid inside the tube.

The level transducer is configured with a pair of 2-KΩ resistors to form a bridge. The required ac bridge excitation is developed at C1A, configured as a multivibrator. C1 biases Q1, which switches the LT1009's 2.5-V potential through the 100-μF capacitor to provide the ac bridge drive. The bridge differential output ac signal is converted to a current by A1, operating as a Howland current pump. This current, whose polarity reverses as bridge drive polarity switches, is rectified by the diode bridge. Thus, the 0.03-μF capacitor receives unipolar charge. A2, running at a differential gain of 2, senses the voltage across the capacitor and presents its single-ended output to C1B. When the voltage across the 0.03-μF capacitor becomes high enough, C1B's output becomes high, turning on the paralleled sections of the LTC1043 switch. This discharges the capacitor. The 47-pF capacitor provides enough ac feedback around C1B to allow a complete zero reset for the capacitor. When the ac feedback ceases, C1B's output decreases and the LTC1043 switch goes off. The 0.03-μF unit again receives constant current charging and the entire cycle repeats. The frequency of this oscillation is determined by the magnitude of the constant current delivered to the bridge-capacitor configuration. This current's magnitude is determined by the transducer bridge's offset, which is level related.

109

Time-Delay Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

ELECTRONIC TIME DELAY

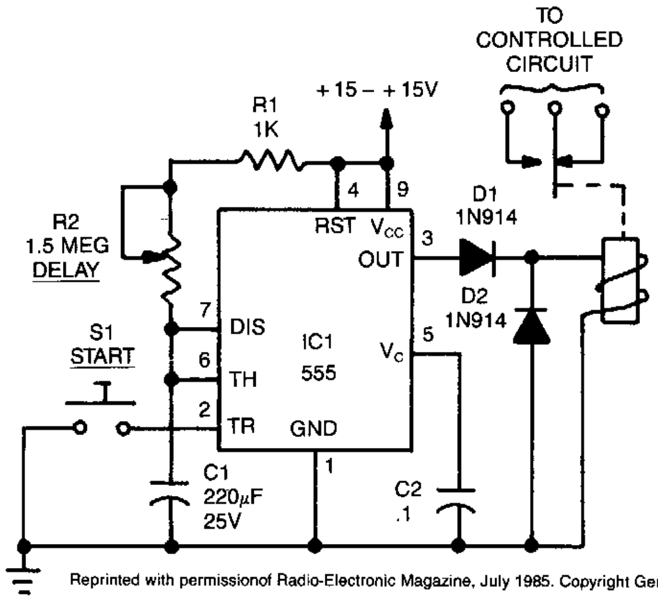
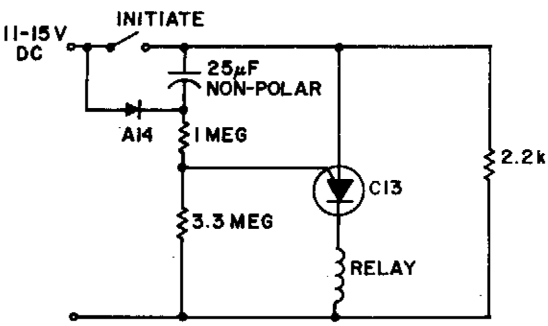


Fig. 109-1

The time delay, T, in seconds is: $T = 1.1 \times C1 \times (R1 + R2)$. The resistances are in megohms and capacitances in microfarads. The sum of R1 and R2 should not be less than 1000 Ω nor higher than 20 MΩ. Pressing S1 starts the timing cycle. A low-going pulse, instead of S1 can also be used to initiate the timing cycle. With the values shown and allowing for the tolerances of the 200-μF capacitor, the delay will range from 4 minutes and 50 seconds to 7 minutes and 26 seconds. The output terminal, pin 3, of 555, is normally low and switches high during the timing cycle. The output can either sink or source currents up to 200 mA. Therefore, a load such as a relay coil can be connected between pin 3 and V_{CC} or between pin 3 and ground, depending on circuit requirements. When the relay is connected between pin 3 and ground, it is normally de-energized so it is energized only during the timing cycle. Connecting the relay to ground will save power and allow the IC to run cool.

TIMING THRESHOLD AND LOAD DRIVER



TIME DELAY 40–60 SECS WITH COMPONENTS SHOWN

GE

Fig. 109-2

Power is applied to the circuit with the initiate switch open. The 25-μF capacitor charges through the A14, or equivalent, diode and 2.2-KΩ resistor to full supply voltage. When the initiate switch is closed, the low side of the capacitor is suddenly raised to +12 V. This raises the diode side of the capacitor to approximately +24 V. The capacitor immediately begins discharging through the series-connected 1 and 3.3-MΩ resistors. Eventually, the C13 gate becomes forward biased, the device turns, and it applies power to the relay. The delay is virtually independent of supply voltage.

SIMPLE TIME DELAY

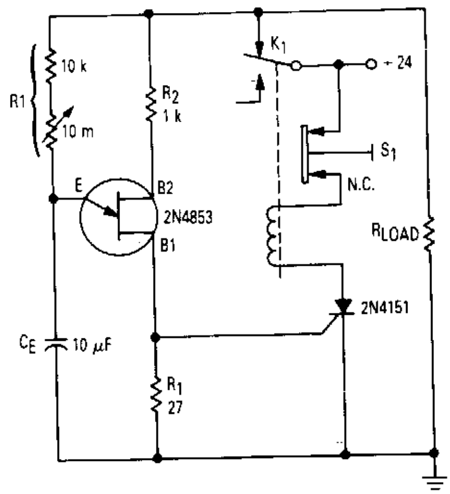


Fig. 109-3

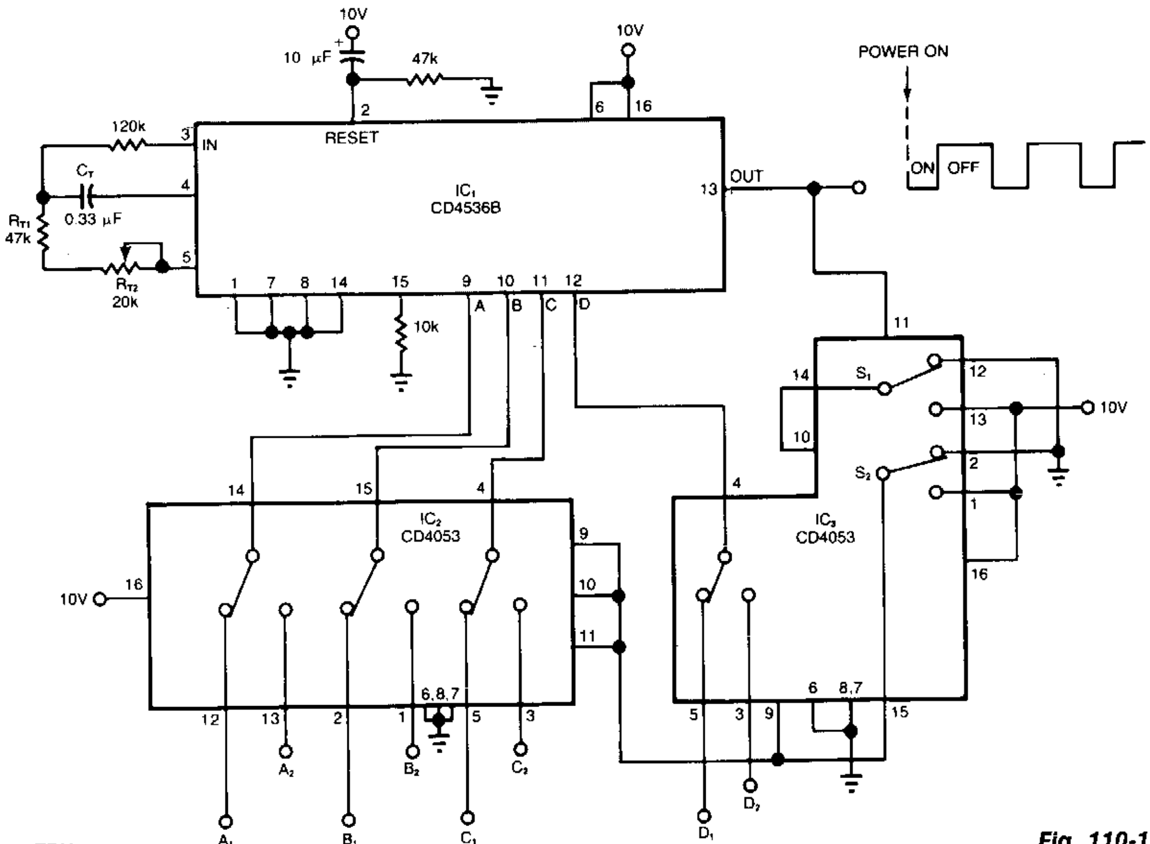
After the first cycle, the relay will normally be energized. When normally closed pushbutton S1 is activated, the SCR turns off, the relay is de-energized, and power is applied to the relaxation oscillator and the load. After a time delay varying from less than a second to approximately 2.5 minutes, as determined by the setting of the 10-MΩ potentiometer, the unijuction will fire and turn on the SCR. The relay will energize until power is removed from the oscillator and the load, and will stay energized until button S1 is pushed again. The UJT trigger output from base 1 directly drives the gate of the SCR. However, where isolation between the UJT trigger or any other type of trigger and the thyristor power circuit is required, then a simple pulse transformer, interfacing the two elements, will suffice.

110

Timers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

SEQUENTIAL TIMER

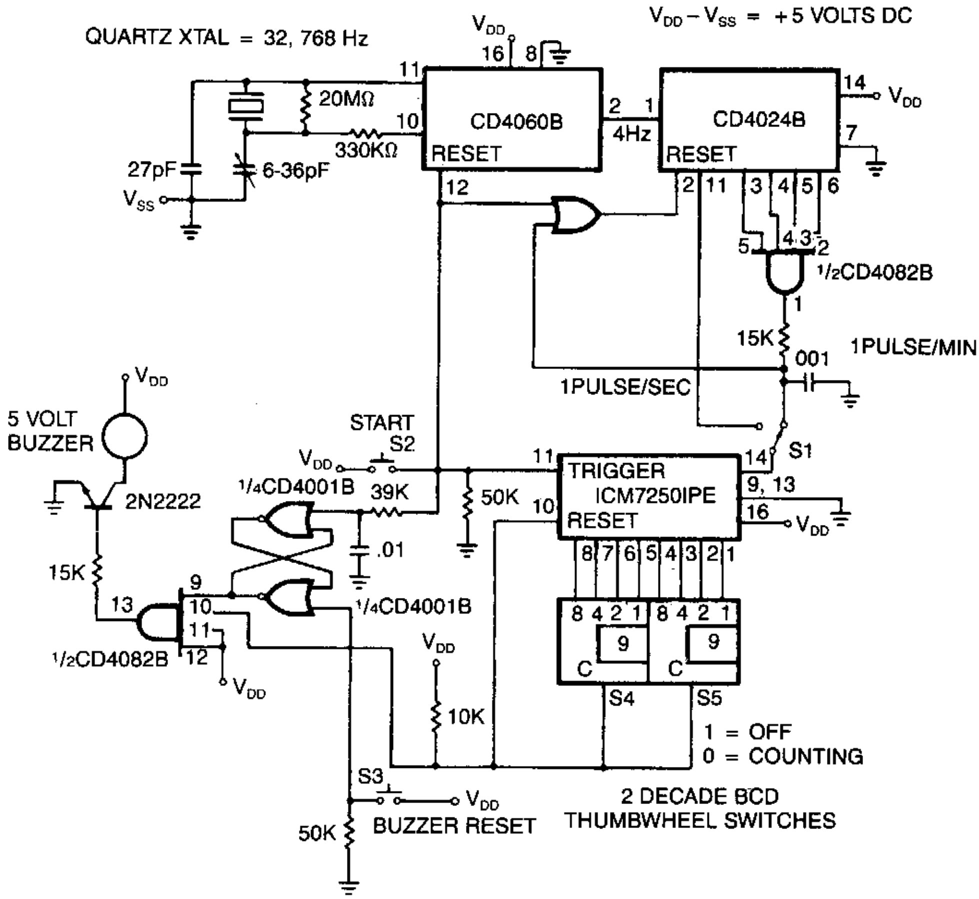


EDN

Fig. 110-1

The timer circuit shown gives independent control of the output's on and off intervals, which can range from 0.055 seconds to 30 minutes, relatively unaffected by power-line transients. IC1 is a CMOS programmable-timer chip that includes 24 ripple-binary counter stages; the first eight are bypassed when logic 1 is applied to pin 6. Then, a 4-bit input code at pins A, B, C, and D connects one of the 16 remaining stages to the output at pin 13. The chip includes an oscillator whose timing components are C_T, R_{T1}, and R_{T2}. For this example, you adjust R_{T2} for an internal period T_{IN} of 54.9 ms (18.2 Hz). Then, the output on or off interval is: $T_{OUT} = T_{IN}2^{N-1}$, where N is the number of counter stages in the internal divider chain (See Fig. 110-3). IC2 and IC3 are CMOS triple-spdt analog switches that connect one BCD code ($A1-D1$) for the on interval and another ($A2-D2$) for the off interval. You can apply the codes using manual toggle switches or programmable latches. When power is first applied, the switches are in the positions shown, which applies $A1-D1$ to IC1 and generates the on interval. When the output changes state, all the switches change position and initiate the off interval by applying $A2-D2$ to IC1. The cycle then repeats. To eliminate race conditions, switches S1 and S2 of IC3 operate in sequence before the remaining four switches operate in parallel. To start the output sequence with an off instead of an on interval, connect a power-on-set signal at pin 1 instead of the power-on-reset signal at pin 2.

CMOS PRECISION PROGRAMMABLE LABORATORY TIMER

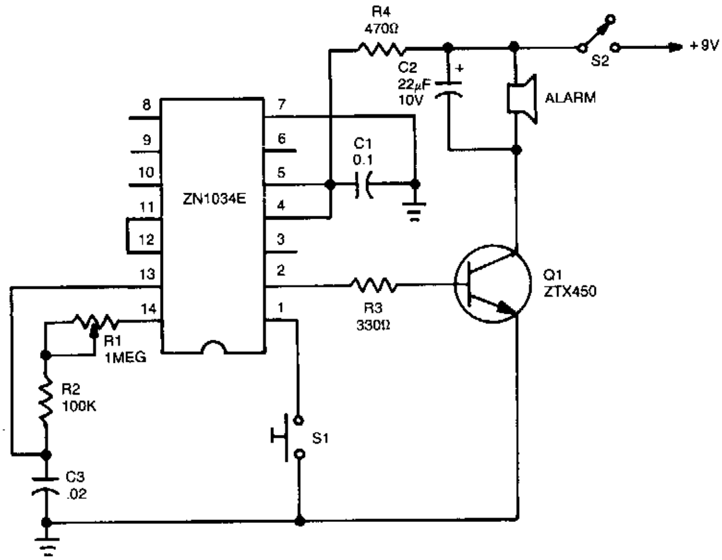


INTERSIL

Fig. 110-2

The time base is first selected with S1 set for seconds or minutes, then units 0 – 99 are selected on the two thumbwheel switches S4 and S5. Finally, switch S2 is depressed to start the timer. Simultaneously, the quartz crystal-controlled divider circuits are reset, the ICM7250 is triggered and counting begins. The ICM7250 counts until the preprogrammed value is reached, then, the value is reset, pin 10 of the CD4082B is enabled, and the buzzer is turned on. Pressing S3 turns the buzzer off.

LONG-TIME TIMER



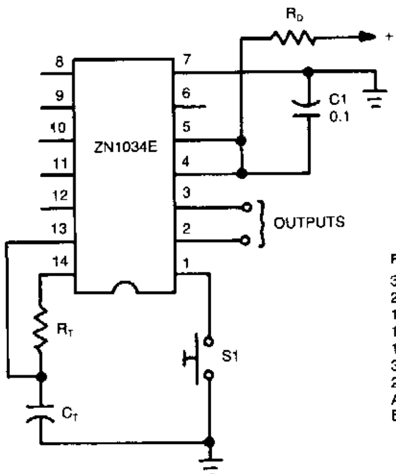


TABLE 1—R_T AND C_T VALUES

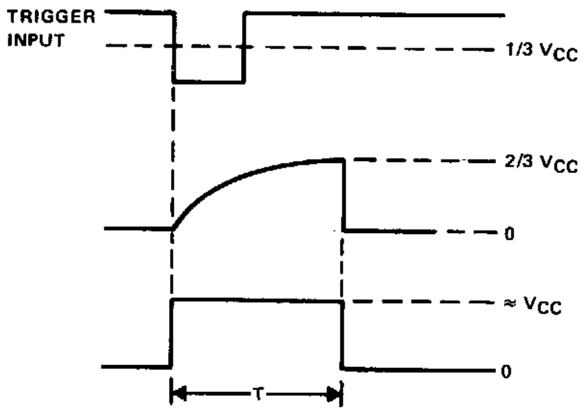
Timing elements		Timed period	
R_T (ohms)	C_T (μ)F	A	B
39K	0.01	1 sec	2.92 sec
220K	0.1	1 min	2.75 min
100K	1.0	5 min	12.5 min
1.2 Meg	1.0	55 min	2.5 hrs.
1.2 Meg	10	9.1 hrs	25 hrs
3.3 Meg	10	1 day	2.8 days
2.2 Meg	100	1 week	19 days

A) Pins 11 and 12 tied together
B) 300K resistor connected between pins 11 and 12

Fig. 110-3

When used as a stand-alone device, ZN1034E from Ferranti can provide timed intervals ranging from 1 second to 19 days, although the rc time constant is only 220 seconds. The ZN1034E includes an internal voltage regulator, an oscillator, and a 12-stage binary counter. The total delay time provided by the counter is 4095 times the oscillator period. The control logic times-out after 4095 cycles of the oscillator, and delivers high and low output pulses at pins 2 and 3. The output at pin 3 is normally high and decreases at the end of the timed interval. The complementary output at pin 2 is normally low and becomes high at the end of the timed interval. The timing period is initiated by momentarily grounding pin 1. Timing resistor R_T consists of two resistors, R1 and R2, in series. Because R1 has a fixed value of 100 KΩ, the total range of R_T is 100 K to 1.1 MΩ.

ONE-SHOT TIMER



NE555 One-Shot Timing Diagram

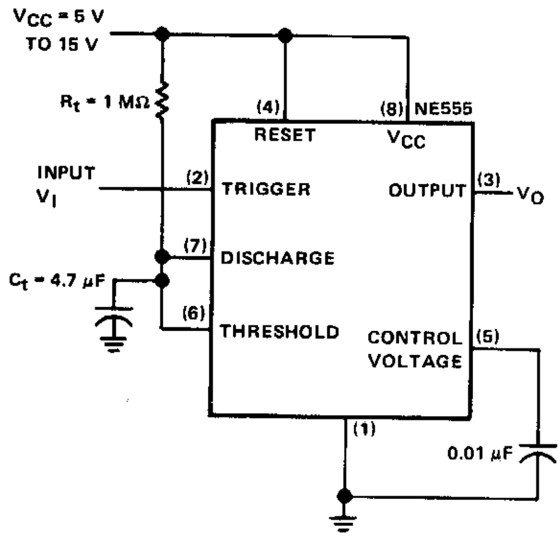
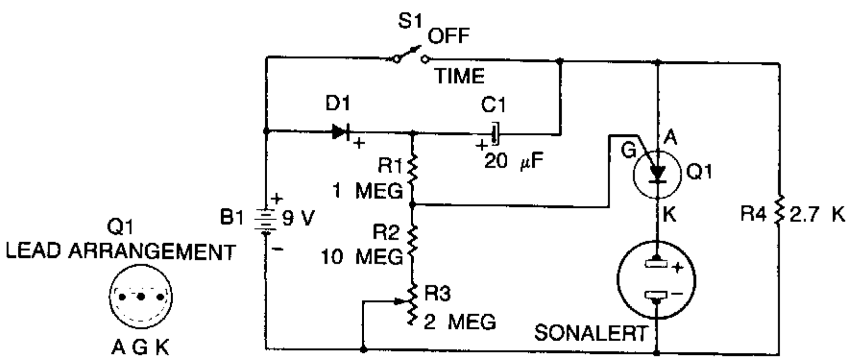


Fig. 110-4

This simple circuit consists of only two timing components R_T and C_T, the NE555, and bypass capacitor C2. While not essential for operation, C2 is recommended for noise immunity. During standby, the trigger input terminal is held higher than $1/3\ V_{CC}$ and the output is low. When a trigger pulse appears with a level less than $1/3\ V_{CC}$, the timer is triggered and the timing cycle starts. The output rises to a high level near V_{CC}, and at the same time, C_T begins to charge toward V_{CC}. When the C_T voltage crosses $2/3\ V_{CC}$, the timing period ends with the output falling to zero, and the circuit is ready for another input trigger. Because of the internal latching mechanism, the timer will always time out when triggered, regardless of any subsequent noise, such as bounce, on the trigger input. For this reason, the circuit can also be used as a bounceless switch by using a shorter rc time constant. A 100-KΩ resistor for R_T and a 1-μF capacitor for C_T would give a clean, 0.1 s output pulse when used as a bounceless switch.

THREE-MINUTE TIMER



TAB BOOKS

Fig. 110-5

THREE-MINUTE TIMER (*Cont.*)

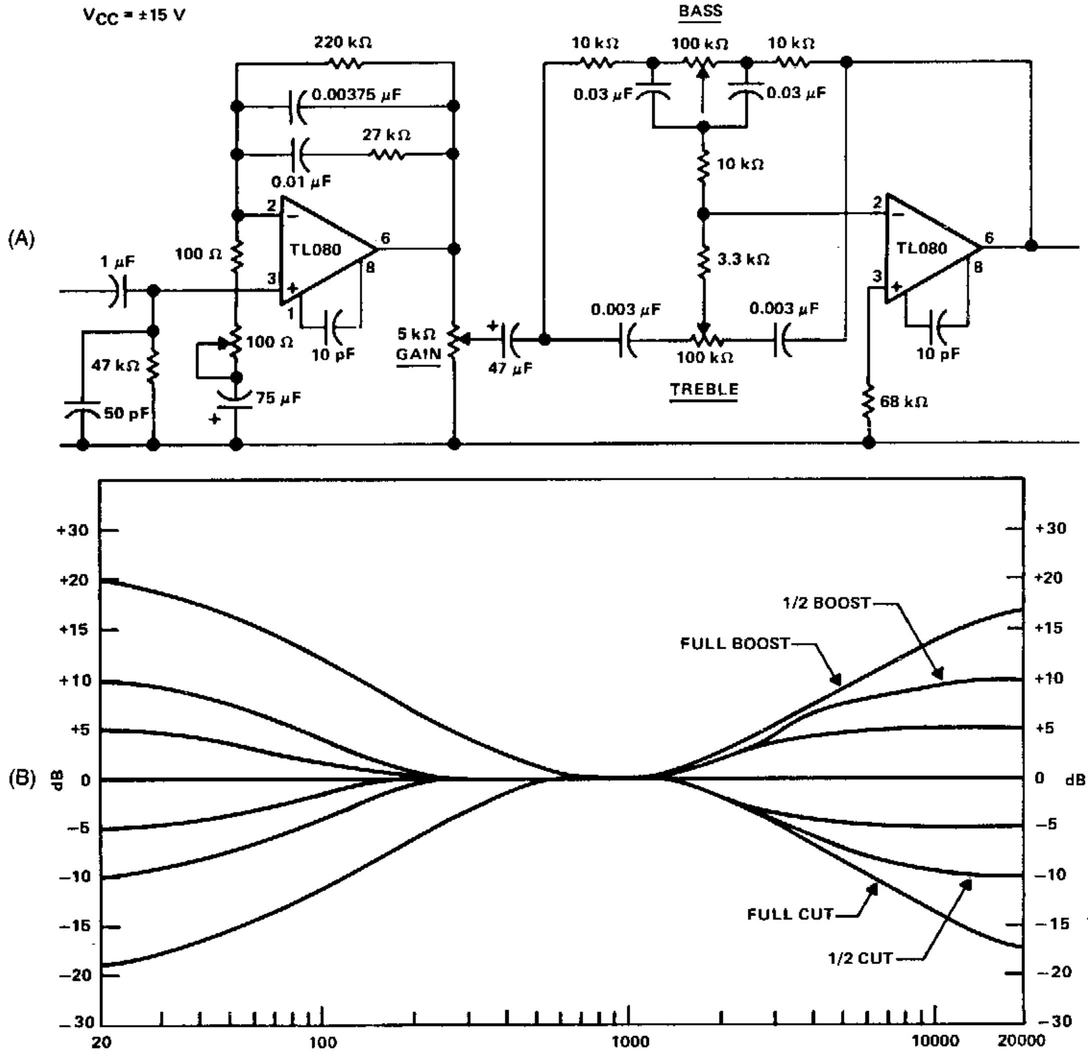
When S1 is off, C1 charges to within 0.5 V of the battery voltage through diode D1 and resistor R4. When S1 is closed, the anode of the PUT rises to the positive supply voltage. The PUT does not conduct, because battery voltage appears in series with the charge stored on C1, which raises the gate of the PUT to a level positive with respect to the anode. The timer relies on the discharge of capacitor C1 through resistors R1, R2, R3,and R4. Once C1 is at zero volts, the PUT will turn on battery voltage to the Sonalert and cause it to sound.

111

Tone Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

IC PREAMPLIFIER/TONE CONTROL

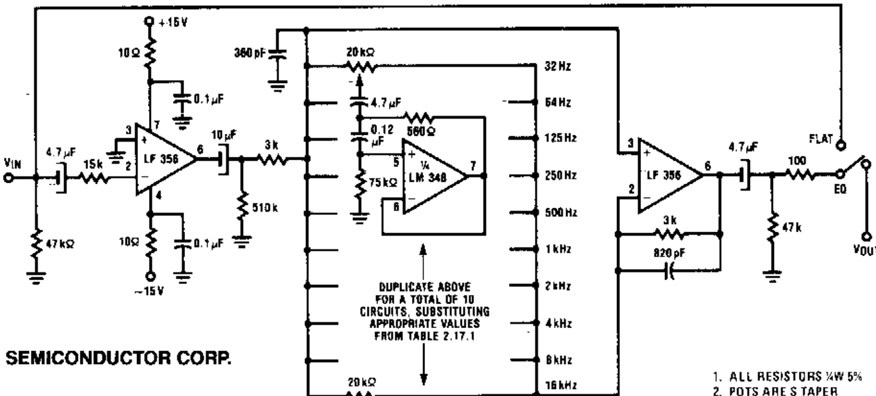


TEXAS INSTRUMENTS

Fig. 111-1

The circuit is a form of the so-called "Americanized" version of the Baxandall negative-feedback tone control. At very low frequencies, the reactance of the capacitor is large enough that they might be considered open circuits, and the gain is controlled by the bass potentiometer. At low to middle frequencies, the reactance of the 0.03-μF capacitors decreases at the rate of 6 dB/octave, and is in parallel with the 200-KΩ potentiometer; so the effective impedance is reduced correspondingly, thereby reducing the gain. This process continues until the 10-KΩ resistors, which are in series with the bass pot, become dominant and the gain levels off at unity. The action of treble circuit is smaller and becomes effective when the reactance of the 0.003-μF capacitors becomes minimal. This complete tone control is in the negative feedback loop of the TL080. Figure B shows the bass and treble tone control response.

TEN-BAND OCTAVE EQUALIZER



NATIONAL SEMICONDUCTOR CORP.

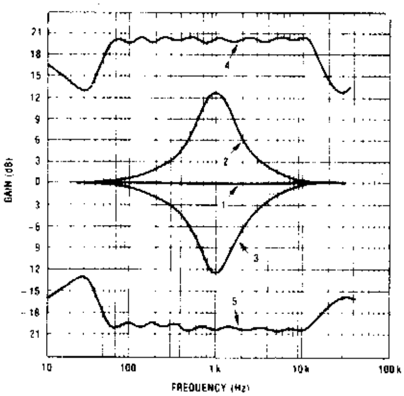
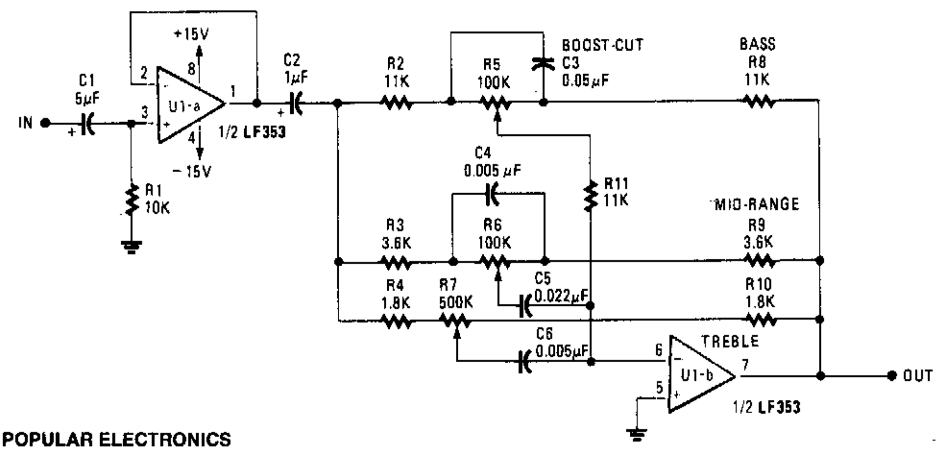


Fig. 111-2

fo(Hz)	C_1	C_2	R_1	R_2
32	0.12μF	4.7μF	75kΩ	560Ω
64	0.056μF	3.3μF	68kΩ	510Ω
125	0.033μF	1.5μF	62kΩ	510Ω
250	0.015μF	0.82μF	68kΩ	470Ω
500	8200pF	0.39μF	62kΩ	470Ω
1k	3900pF	0.22μF	68kΩ	470Ω
2k	2000pF	0.1μF	68kΩ	470Ω
4k	1100pF	0.056μF	62kΩ	470Ω
8k	510pF	0.022μF	68kΩ	510Ω
16k	330pF	0.012μF	51kΩ	510Ω

A series of active rf filters using National LM348IC comprises a ten-band graphic equalizer. C1, C2, R1, and R2 should be at least 10% with 5% preferred tolerances.

THREE-BAND ACTIVE TONE CONTROL



POPULAR ELECTRONICS

Fig. 111-3

WIEN-BRIDGE FILTER

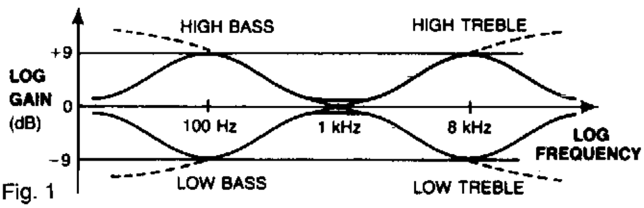


Fig. 1

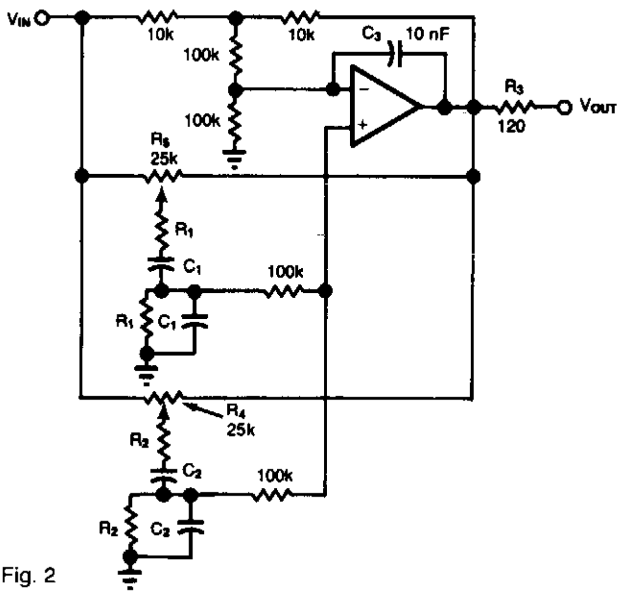


Fig. 2

EDN

Fig. 111-4

Most audio tone controls affect midband gain, and they often create booming or hissing sounds when activated. You can avoid these problems by using a dual Wien-bridge filter to provide independent control of the treble and bass frequencies.

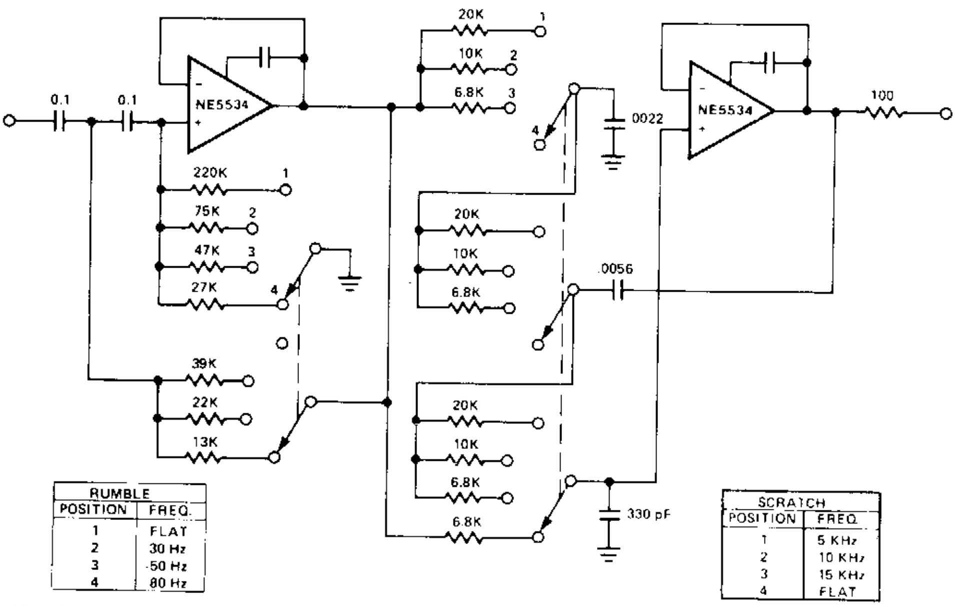
Experiments with equalizers indicate that the optimum center frequencies are about 100 Hz and 8 kHz. Using the relation $f = (2\pi RC)^{-1}$, set the Fig. 1 values accordingly:

100 Hz: R1 = 15 KΩ; C1 = 0.1 μF

8 kHz: R2 = 16 KΩ; C2 = 1.3 nF

R3 and C3 provide stability. You obtain a ± 9 dB variation of treble and bass by adjusting potentiometers R4 and R5, respectively. The filter's frequency response is shown in Fig. 2.

RUMBLE/SCRATCH FILTER



RUMBLE	
POSITION	FREQ.
1	FLAT
2	30 Hz
3	50 Hz
4	80 Hz

SCRATCH	
POSITION	FREQ.
1	5 KHz
2	10 KHz
3	15 KHz
4	FLAT

All resistor values are in ohms.

SIGNETICS ***Fig. 111-5***

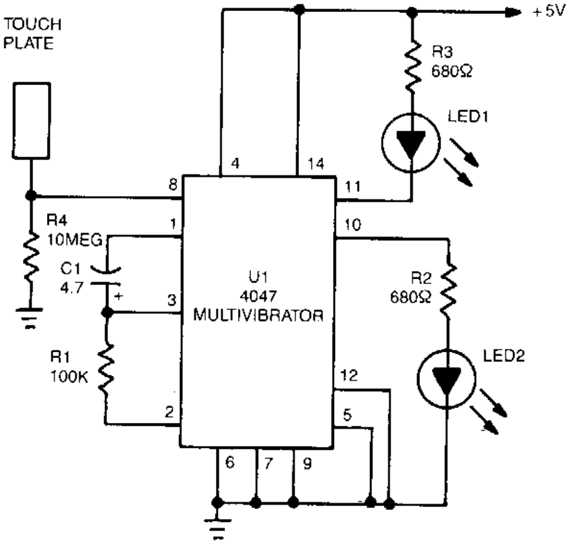
This is a variable bandpass amplifier with adjustable low- and high-frequency cutoffs.

112

Touch-Switch Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

NEGATIVE-TRIGGERED TOUCH CIRCUIT



POPULAR ELECTRONICS

Fig. 112-1

The 4047 is configured as a monostable multivibrator circuit or one shot that is set up to trigger on a negative-transition of the signal applied to its pin 6 input. The multivibrator's on time is determined by the values of R1 and C1. Although R1 is shown to be a 100-K unit, its value can be anything between 10 K and 1 MΩ. Capacitor C1 can be a nonpolarized capacitor with any practical value above 100 pF. By making R4's value extremely high, the circuit can be used as a touch-triggered one-shot multivibrator. If the value of R4 is reduced to a much lower value, such as 10 KΩ, the circuit can be triggered with a negative pulse through 0.1-μF capacitor connected to pin 6. With a 100-KΩ resistor for R1, and a 4.7-μF electrolytic capacitor for C1, the circuit's on time is about 0.6 second. When R1 is increased to 470 KΩ, the on time of the circuit is increased to over 6 seconds.

POSITIVE-TRIGGERED TOUCH CIRCUIT

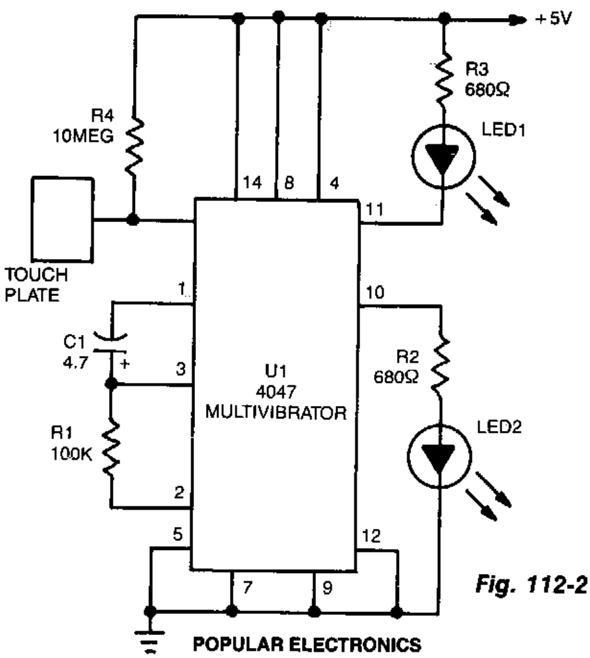
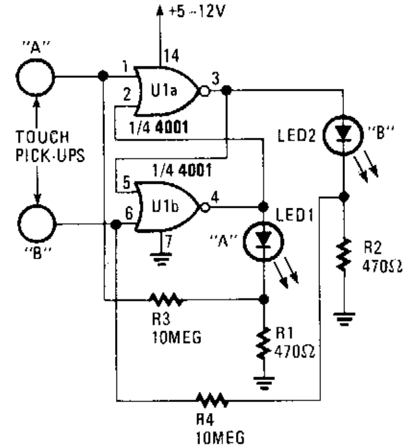


Fig. 112-2

POPULAR ELECTRONICS

LED1 and LED2 indicators turn on and remain on, each time the circuit is triggered. During the timing cycle, U1's Q output at pin 10 becomes positive when the Q output at pin 11 becomes negative. The two LEDs can be removed and the Q and Q outputs at pins 10 and 11, respectively, can be used to trigger some other circuit.

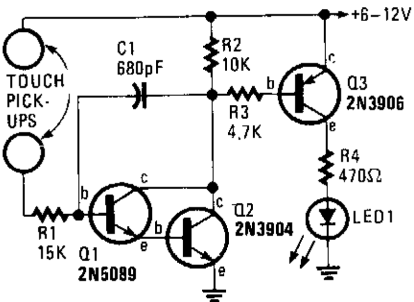
DIGITAL TOUCH ON/OFF SWITCH



HANDS-ON ELECTRONICS *Fig. 112-3*

Only one LED can be on when the circuit is at rest. Which LED is illuminated is determined by the touch pick-up that last had human contact. Pickup terminal A controls the on condition of LED1, and terminal B controls the on condition of LED2. A 4001 quad two-input NOR gate is connected in an anti-bounce latching circuit that is activated by touching a pickup.

TWO-TERMINAL TOUCH SWITCH



HANDS-ON ELECTRONICS *Fig. 112-4*

This circuit requires the bridging of two circuits to activate the electronic switch. That circuit does not require a 60-Hz field to operate and can be battery or ac powered. The two-pickup terminals can be made from most any clean metal; they should be about the size of a penny. The input circuitry of the two-terminal touch switch is a high-gain Darlington amplifier that multiplies the small bridging current to a value of sufficient magnitude to turn on Q3, supplying power to LED1. If a quick on and off switching time is desired, the value of C1 should be very small; if a long on-time period is required, the value of C1 can be increased.

TOUCH ON/OFF ELECTRONIC SWITCH

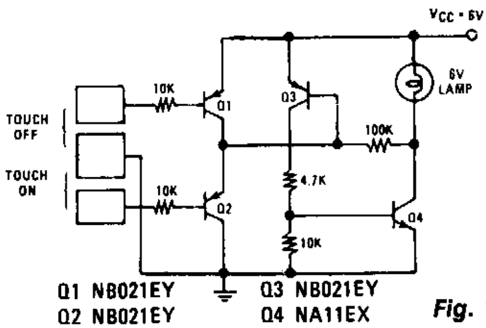
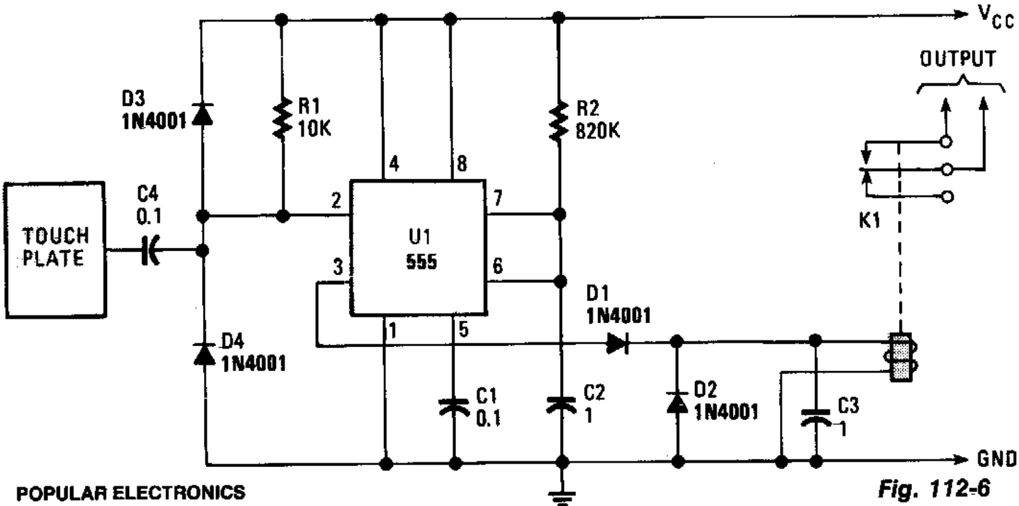


Fig. 112-5

Transistors Q1 and Q2 control latch Q3 and Q4 to switch on the lamp. A high resistance from touching the electrode biases Q1 or Q2 on, setting or resetting the latch.

NATIONAL SEMICONDUCTOR CORP.

LINE-HUM TOUCH SWITCH



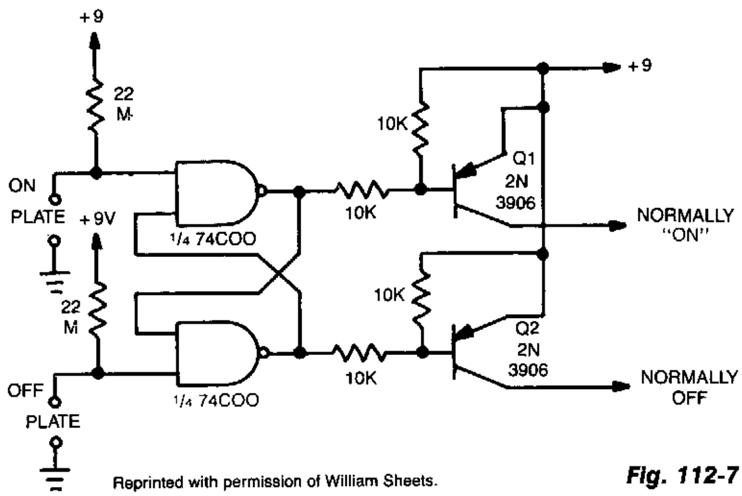
POPULAR ELECTRONICS

Fig. 112-6

The monostable period is set for about 1 second, as is the usual case. The induced line hum comes through C2, providing a continuous string of trigger pulses. The output becomes low for about 10 ms per second as the monostable times out and then retriggers. Diode D1 and capacitor C3 buffer the relay so it doesn't chatter on those 10-ms pulses. Resistor R2 sets the sensitivity.

The relay energizes when the plate is touched and de-energizes, up to one second after the finger is removed. The delay is a function of when the monostable last retriggered.

TOUCH SWITCH

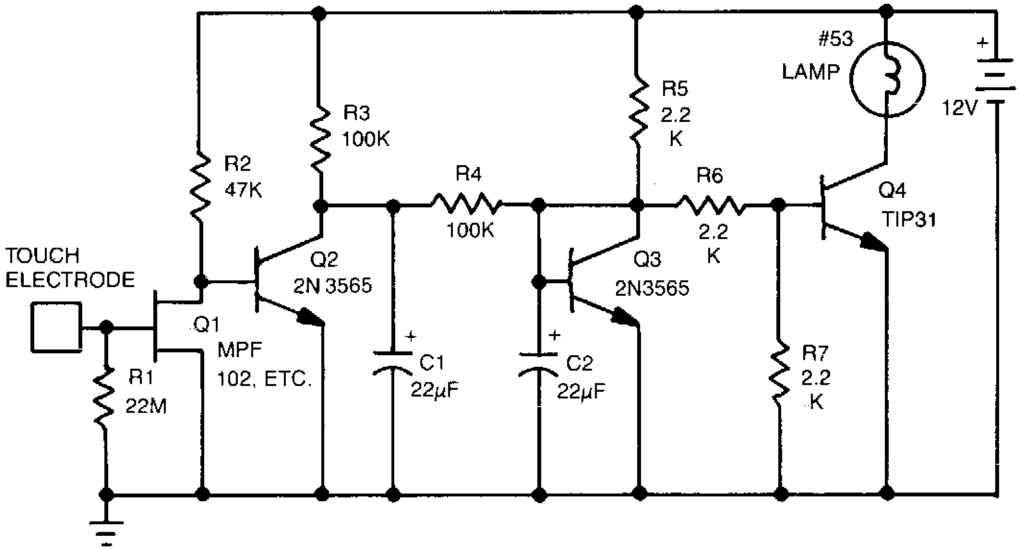


Reprinted with permission of William Sheets.

Fig. 112-7

When the plate is touched, the gate input becomes low, changing the state of the latch. Q1 and Q2 give alternate N-on—N-off outputs.

TOUCH SWITCH

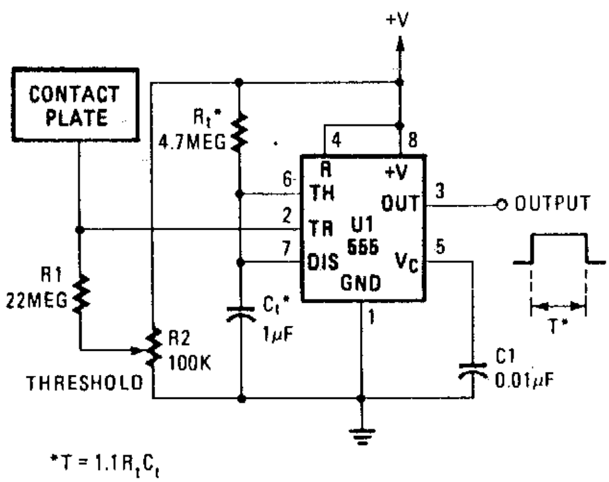


Reprinted with permission of William Sheets.

Fig. 112-8

This touch-actuated switch stays on as long as you keep your finger on the touch plate. R1 sets the input impedance to a high 22 MΩ. Q1 picks up stray signals coupled through your body to the touch plate and amplifies them to turn on Q2, which turns on lamp drivers Q3 and Q4. Lamp I1 is any small 12-V lamp, such as a No. 53—12 V 120 mA. R4 and C1 add a small amount of hysteresis (delay) to keep the light from constantly flickering. A relay can be used for I1.

TOUCH SWITCH



$^*T = 1.1 R_t C_t$

(CHOOSE R_t AND C_t FOR PULSE WIDTH GREATER THAN ANTICIPATED CONTACT TIME.)

HANDS-ON ELECTRONICS

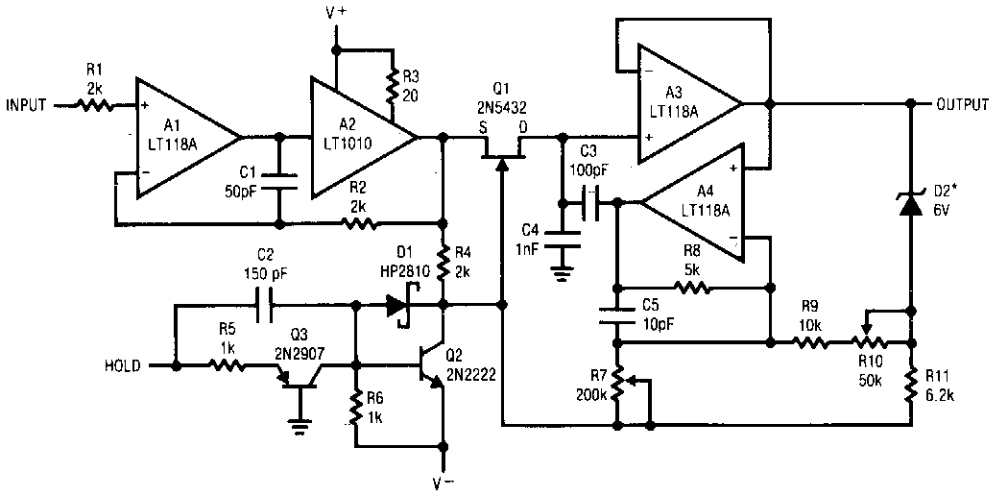
Fig. 112-9

113

Tracking Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

TRACK AND HOLD

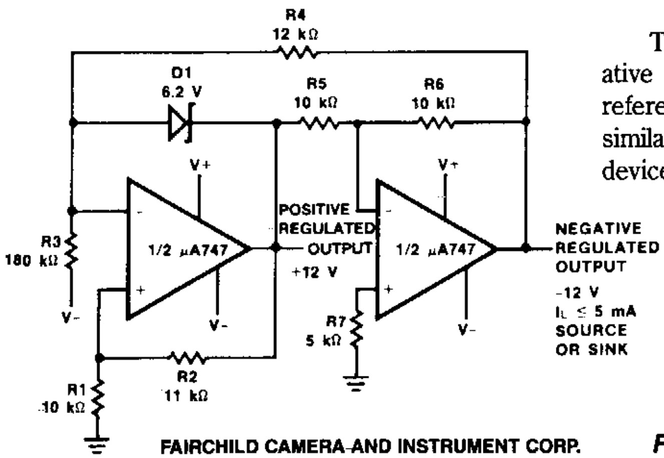


LINEAR TECHNOLOGY CORP. *2N2369 EMITTER BASE JUNCTION ***Fig. 113-1***

The 5-MHz track and hold shown here has a 400-kHz power bandwidth driving ± 10 V. A buffered input follower drives the hold capacitor, C4, through Q1, a low resistance FET switch. The positive hold command is supplied by TTL logic, with Q3 level shifting to the switch driver, Q2. The output is buffered by A3. When the gate is driven to V− for hold, it pulls the charge out of the hold capacitor. A compensating charge is put into the hold capacitor through C3. The step into hold is made independent of the input level with R7, and adjusted to zero with R10.

Since internal dissipation can be quite high when driving fast signals into a capacitive load, using a buffer in a power package is recommended. Raising the buffer quiescent current to 40 mA with R3 improves frequency response.

POSITIVE AND NEGATIVE VOLTAGE REFERENCE TRACKER



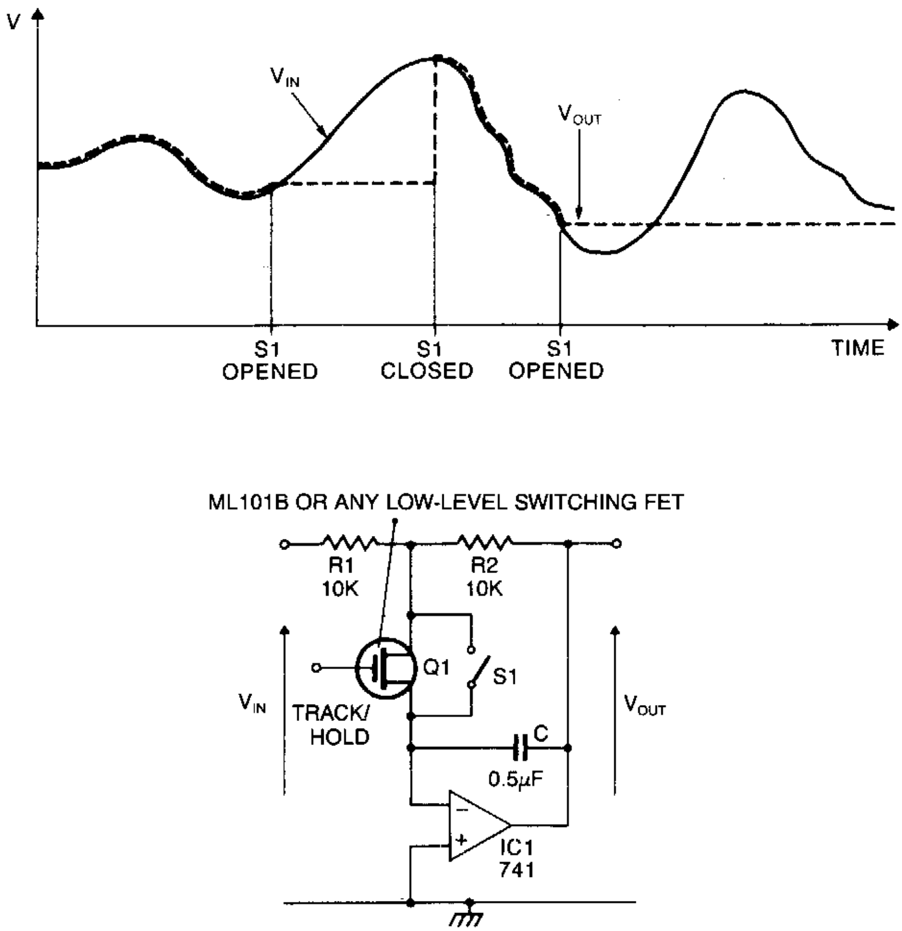
This reference uses an op amp to derive a negative output voltage that tracks with the positive reference voltage. A μA747 dual op amp, or any similar device such as an LM1458 or two μA741 devices, can be used.

$$\text{Positive Output} = V_{D1} \times \frac{R1 + R2}{R2}$$

$$\text{Negative Output} = -\text{Positive Output} \times \frac{R6}{R5}$$

FAIRCHILD CAMERA AND INSTRUMENT CORP. ***Fig. 113-2***

SIGNAL TRACK AND HOLD



ELECTRONICS TODAY INTERNATIONAL

Fig. 113-3

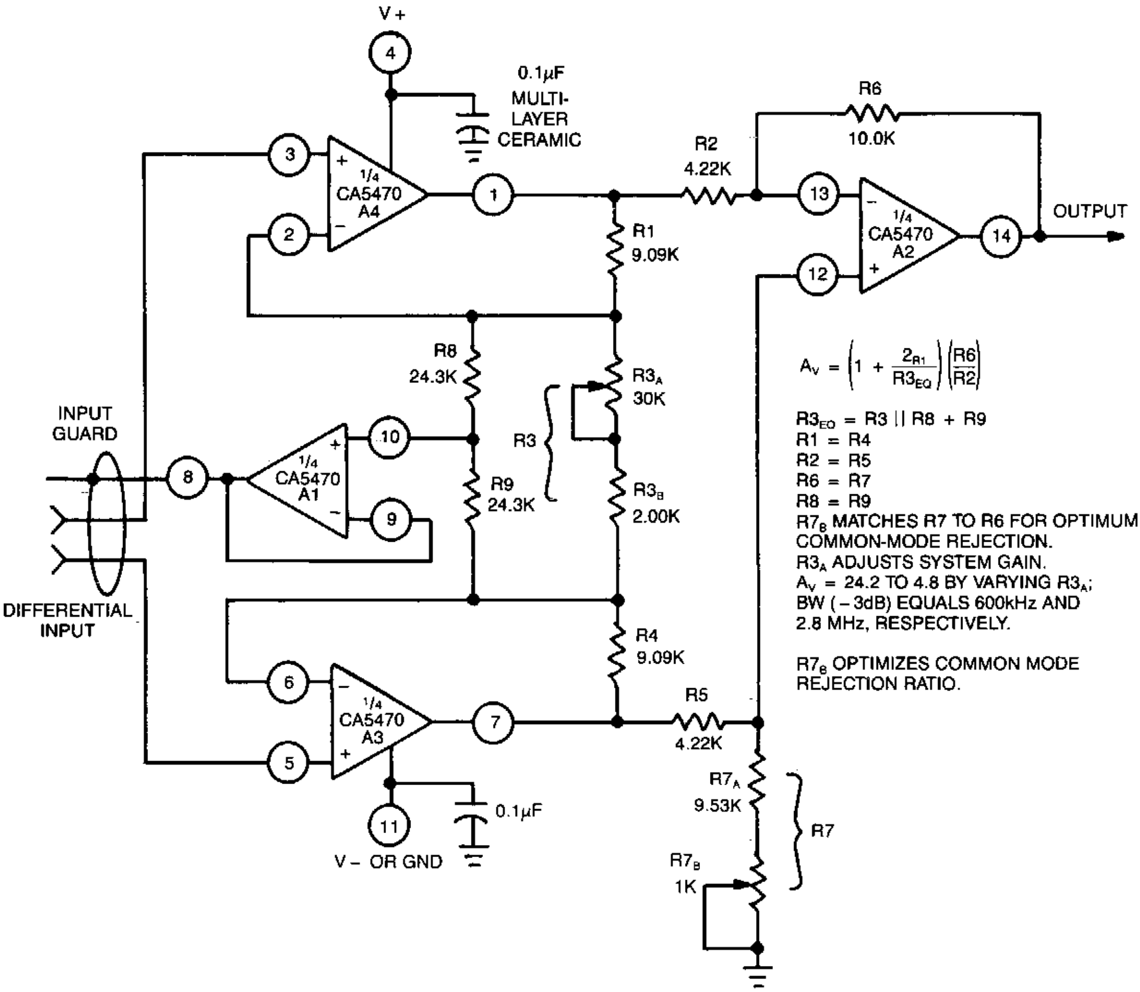
When the switch is closed or the FET is conducting, the circuit behaves as an inverting amplifier with a gain of R2/R1. Since as the inverting terminal of the op amp is a virtual ground, the capacitor is kept charged to the output voltage by the op amp. When the switch is opened and the FET is nonconducting, the voltage at the output is held constant by the capacitor, the current demands of the next stage are met by the op amp. The value of C should be chosen so that its impedance at the operating frequency is large compared to R1 and R2.

114

Transducer Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

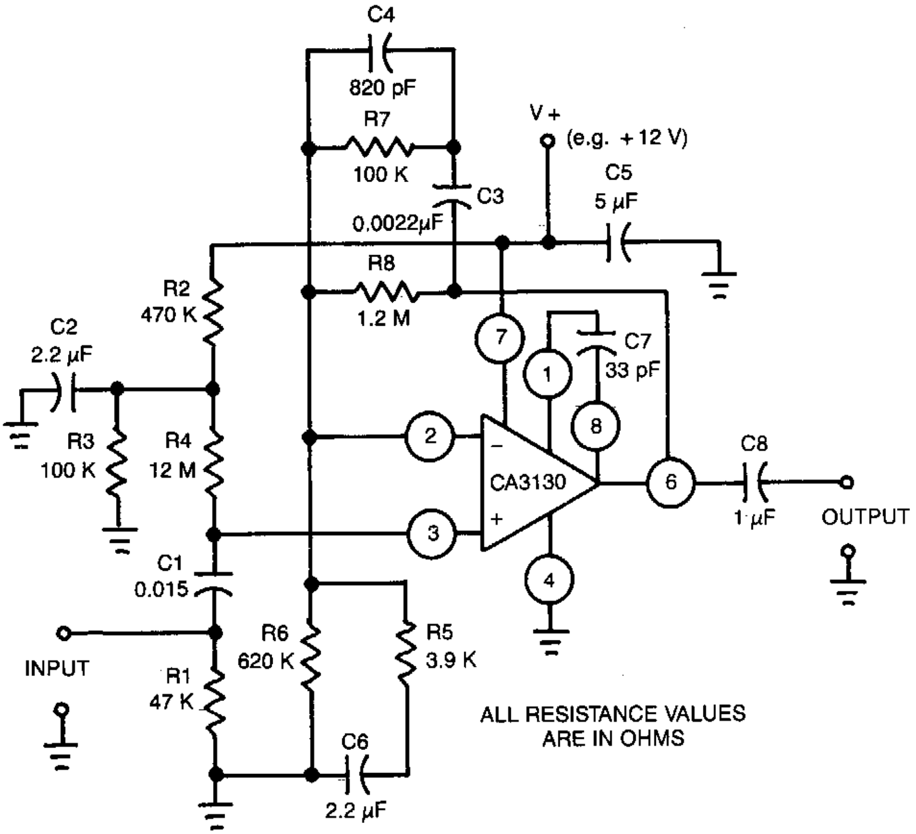
DIFFERENTIAL-TO-SINGLE-ENDED VOLTAGE AMPLIFIER



GE/RCA *Fig. 114-1*

This circuit uses a CA5470 quad microprocessor BiMOS-E op amp. Amplifiers A1 and A2 are employed as a cross-coupled differential input and differential output preamp stage and A3 provides input guard-banding. Amplifier A4 converts the differential outputs of A1 and A2 to a single-ended output.

EQUALIZED PREAMP FOR MAGNETIC PHONOGRAPH CARTRIDGES

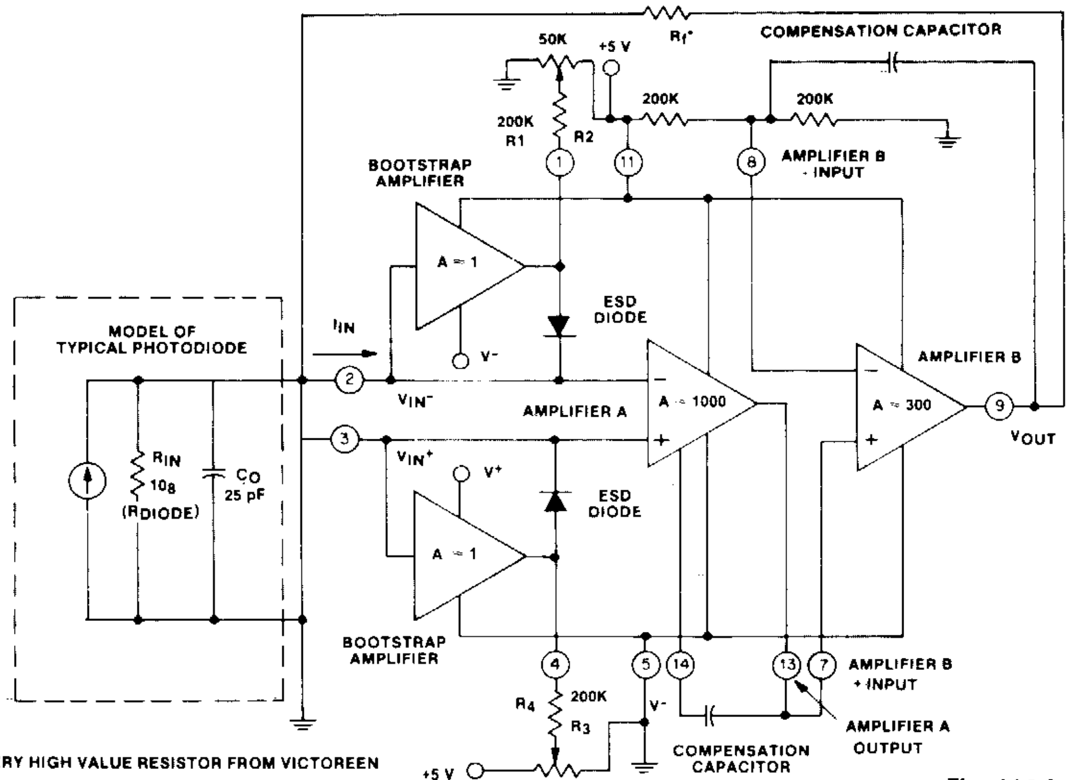


GE/RCA

Fig. 114-2

This circuit uses a CA3130 BiMOS op amp. Amplifier is *equalized* to RIAA playback frequency-response specifications. The circuit is useful as preamplifier following a magnetic tapehead.

PHOTODIODE AMPLIFIER



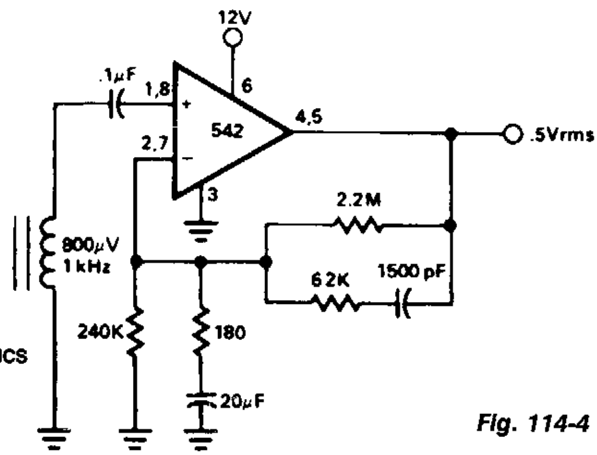
*VERY HIGH VALUE RESISTOR FROM VICTOREEN

GE/RCA

Fig. 114-3

This circuit uses a CA5422 dual BiMOS microprocessor op amp. The bootstrap amplifiers minimize bias currents while maintaining electrostatic discharge protection. Additionally, the potentiometers and their associated resistors, R1 through R4, permit the user to trim bias currents to zero.

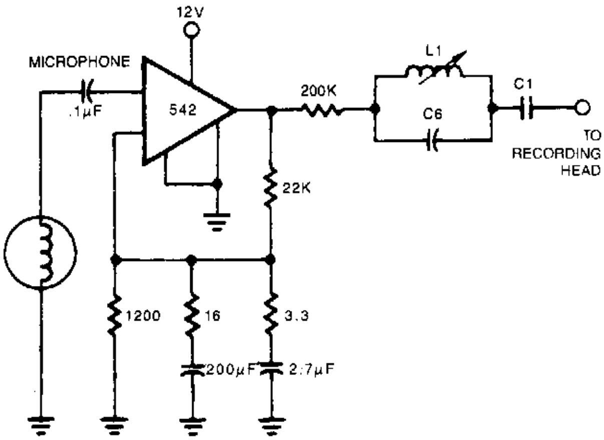
TAPE PLAYBACK AMPLIFIER



SIGNETICS

Fig. 114-4

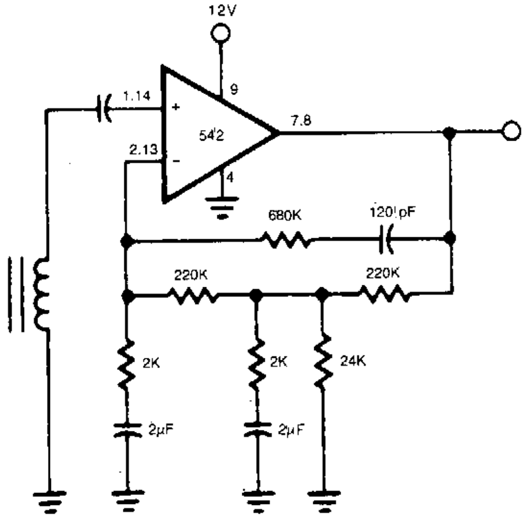
NAB RECORD PREAMPLIFIER



NOTE:
All resistor values are in Ω.

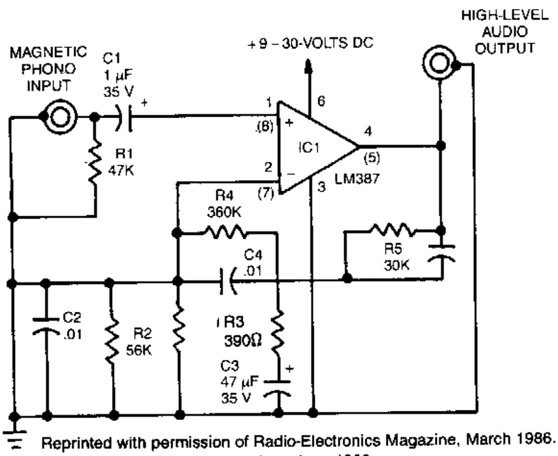
SIGNETICS *Fig. 114-5*

TWO-POLE NAB-TYPE PREAMP



SIGNETICS *Fig. 114-7*

MAGNETIC PHONO PREAMPLIFIER

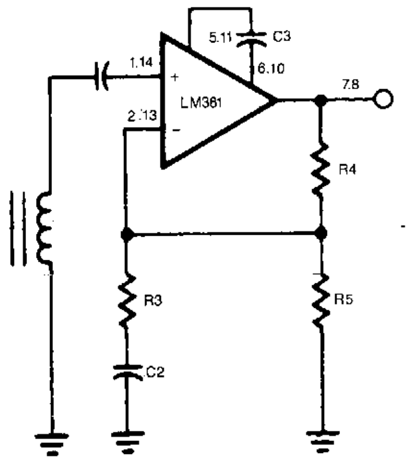


Reprinted with permission of Radio-Electronics Magazine, March 1986.

Fig. 114-6

This simple stereo amplifier uses a National LM387IC. The pin numbers in parentheses are for one channel, and those not in parentheses are for the other channel. The supply voltage can be +9 to +30 Vdc at about 10 mA. The output voltage swing is about $V_{CC}-2$ V pk-pk. The preamp should be able to deliver at least 5 V.

FLAT-RESPONSE TAPE AMPLIFIER



SIGNETICS *Fig. 114-8*

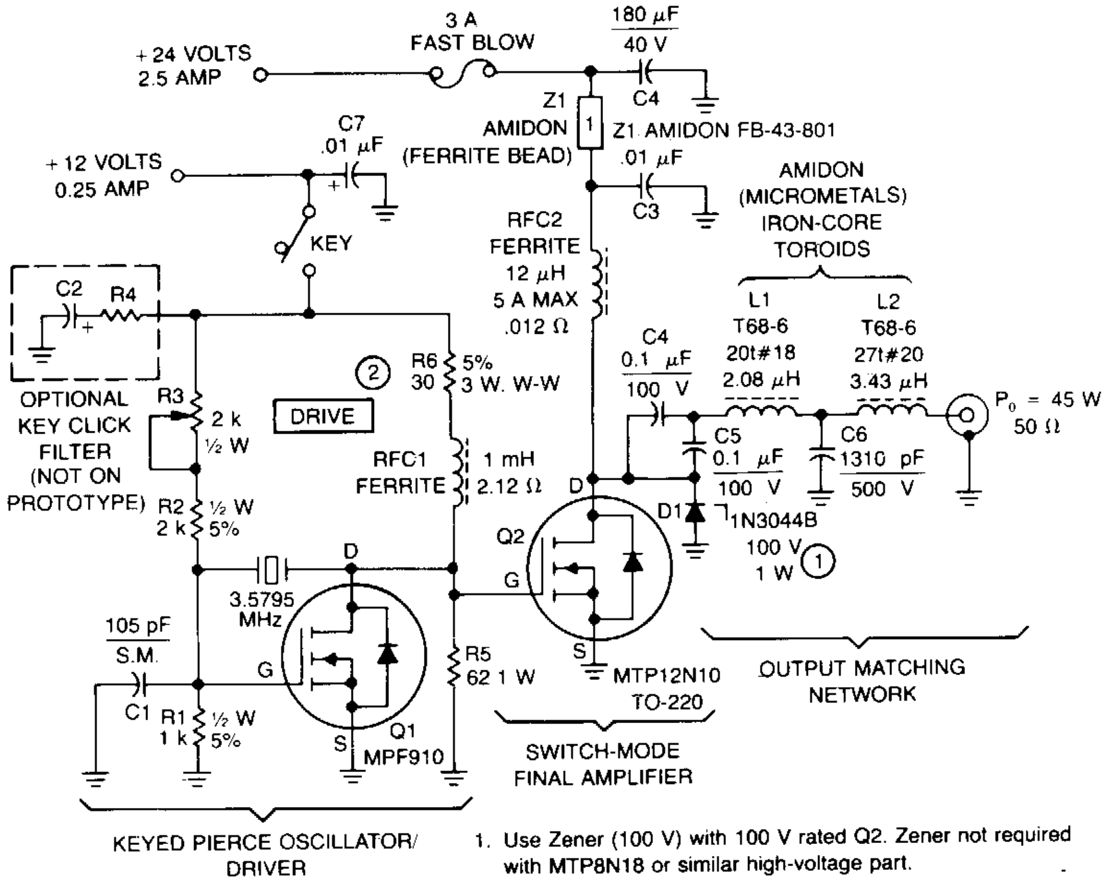
115

Transmitters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

80-M Amateur Radio Transmitter
TV Transmitter
FM Voice Transmitter
1-W CW Transmitter
Low-Cost Half-Duplex Information Transmission Link
FM Snooper
VHF Tone Transmitter
Low-Frequency Transmitter
Wireless FM Microphone
Beacon Transmitter
40-M CW Transmitter
VHF Modulator
Wireless FM Microphone
902-MHz CW Transmitter
One-Transistor FM Transmitter
FM Multiplex Transmitter
QRP CW Transceiver
Wireless FM Microphone

80-M AMATEUR RADIO TRANSMITTER



1. Use Zener (100 V) with 100 V rated Q2. Zener not required with MTP8N18 or similar high-voltage part.
2. Adjust DRIVE for minimum oscillation delay on keying.

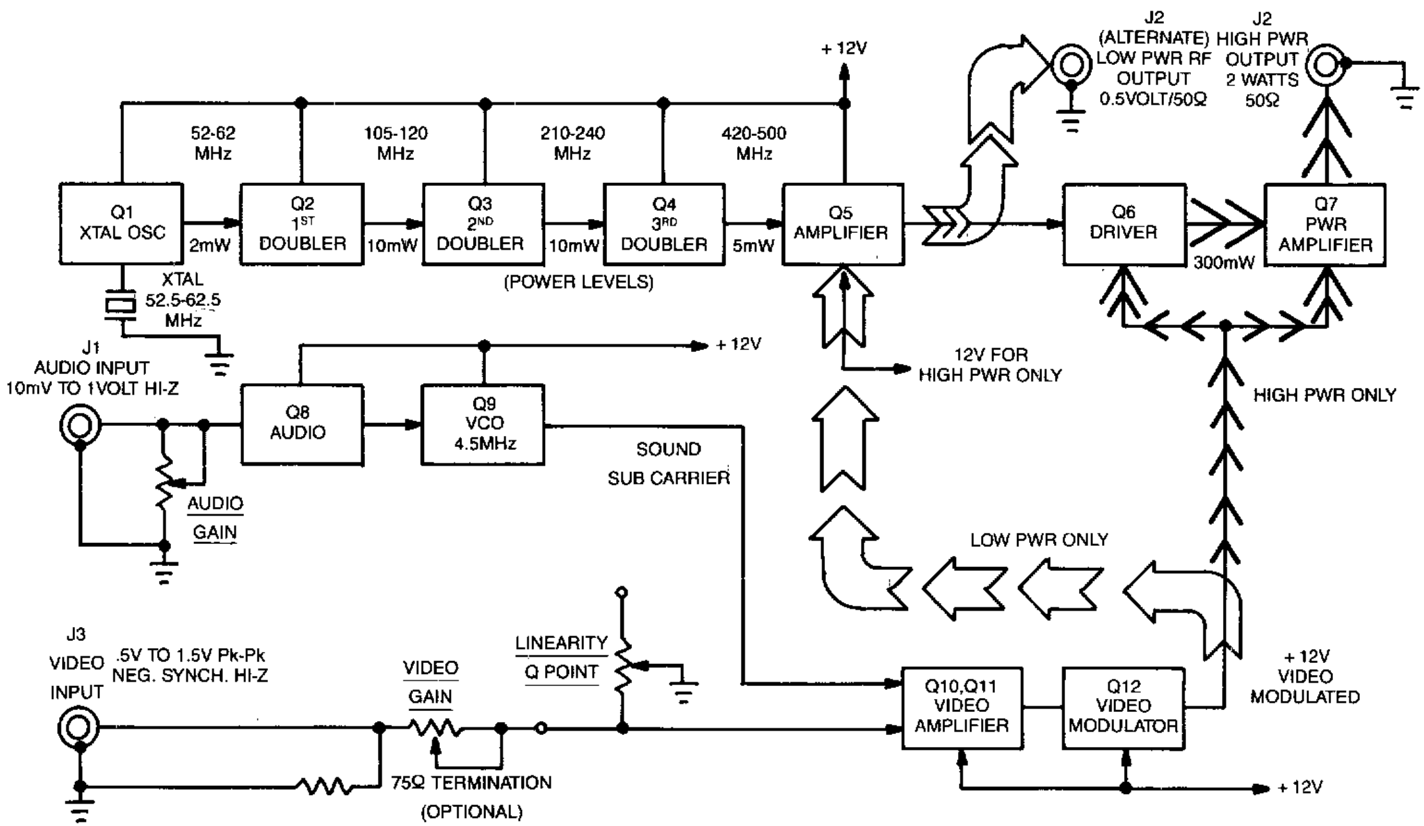
Fig. 115-1

This transmitter consists of a keyed crystal oscillator/driver and a high efficiency final, each with a TMOS Power FET as the active element. The total parts cost less than $20, and no special construction skills or circuit boards are required.

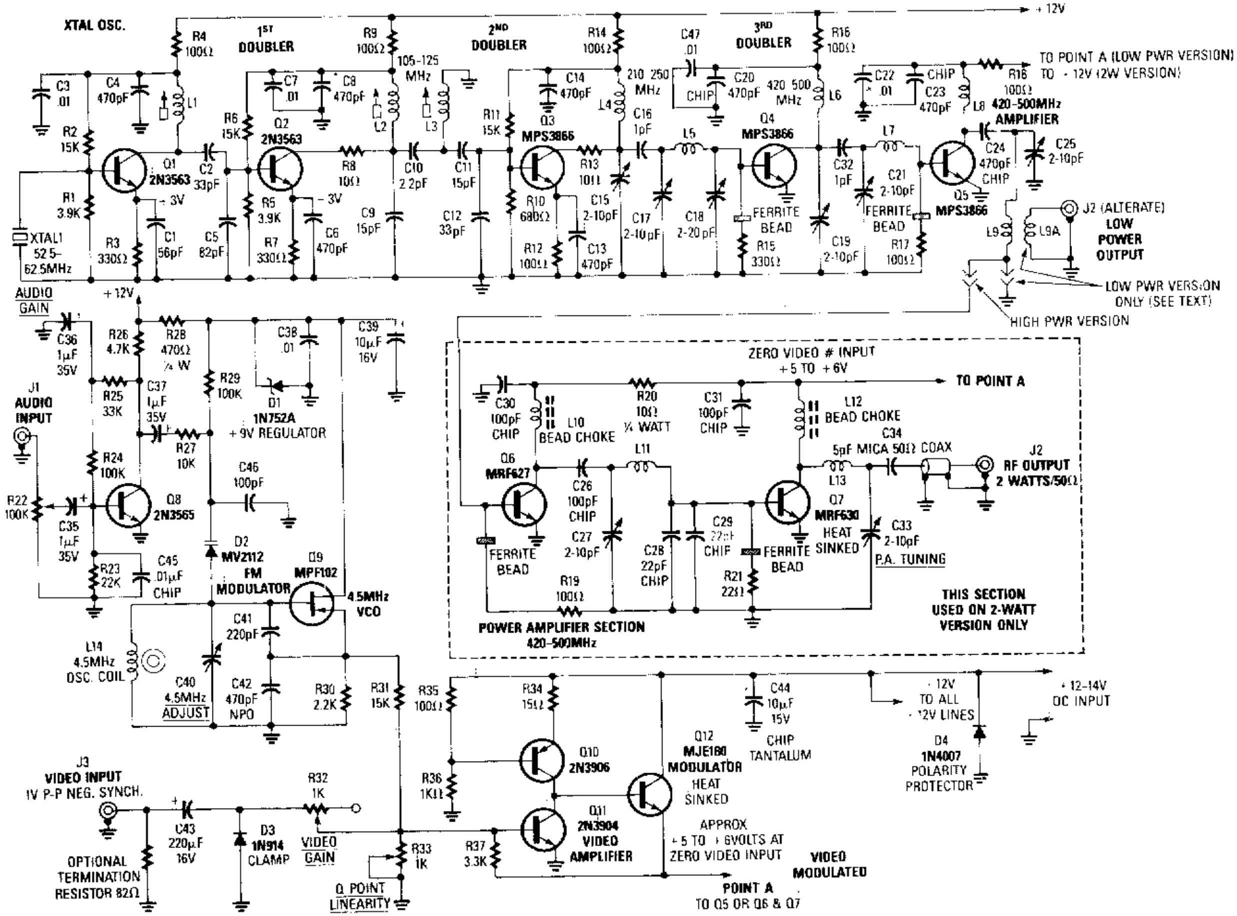
The Pierce oscillator is unique because the high C_{RSS} of the final amplifier power FET, 700 – 1200 pF, is used as part of the capacitive feedback network. In fact, the oscillator will not work without Q2 installed. The MPF910 is a good choice for this circuit because the transistor is capable of driving the final amplifier in a switching mode, while still retaining enough gain for oscillation. To minimize cost, a readily-available color burst TV crystal is used as the frequency-determining element for Q1.

An unusual 84% output efficiency is possible with this transmitter. Such high efficiency is achieved because of the TMOS power FET's characteristics, along with modification of the usual algorithm for determining output matching.

TV TRANSMITTER



This transmitter is capable of two levels of rf power. For low-power wireless video, like in a house or office, where simultaneous monitoring of program material is desirable without cumbersome hookups, 1 – 30 mW is available. For longer ranges up to several miles, as in amateur (ham) TV, security, and surveillance purposes, 2 W into a 50-Ω load is available.

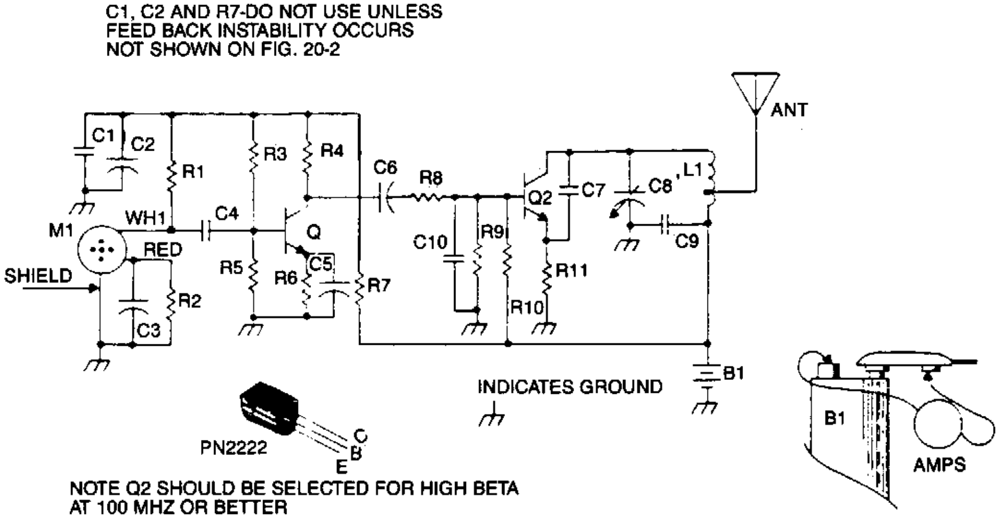


Reprinted with permission from Radio-Electronics Magazine, June 1989. Copyright Gernsback Publications, Inc., 1989.

Fig. 115-2

The video-link transmitter accepts color and B/W video, and audio inputs from VCRs, camcorders, small TV cameras, and microphones. The unit runs on a nominal 12 Vdc and draws 100 mA in the low-power version, or 500 mA in the 2-W version. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

FM VOICE TRANSMITTER

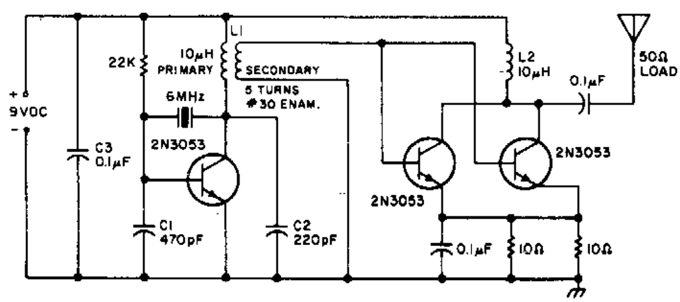


TAB BOOKS

Fig. 115-3

This is a sensitive, mini-powered FM transmitter consisting of an rf oscillator section interfaced with a high-sensitivity wide passband audio amplifier and capacitance microphone with built-in FET that modulates the base of the rf oscillator transistor. The setting of C8 determines the desired operating frequency—in the standard FM broadcast band, tuned to favor the high end up to 110 MHz. Capacitor C7 supplies the necessary feedback voltage developed across R11 in the emitter circuit of Q2, sustaining an oscillating condition. Resistors R9 and R10 provide the necessary bias of the base-emitter junction for proper operation, and capacitor C10 bypasses any rf to ground fed through to the base circuit. C9 provides an rf return path for the tank circuit of L1 and C8, while blocking the dc supply voltage fed to the collector of Q2. The speech voltage developed across R1 by M1 is capacitively coupled by C4 to the base of Q1. A signal voltage developed across R4 is capacity-coupled through C6 to the base of Q2 through R8. R7 and R8, along with C1 and C2, decouple the oscillator and audio circuits.

1-W CW TRANSMITTER



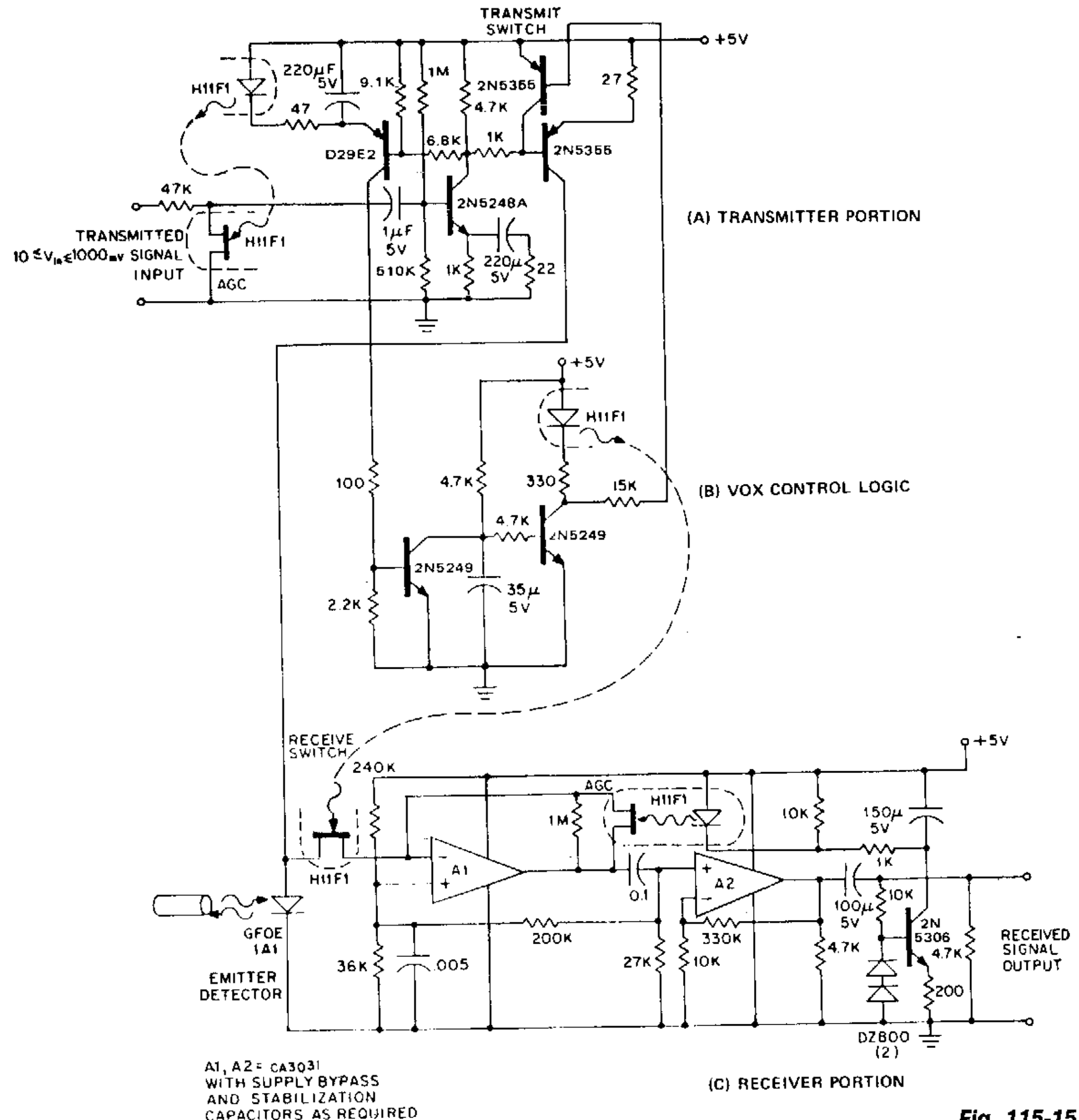
73 AMATEUR RADIO

Fig. 115-4

1-W CW TRANSMITTER (*Cont.*)

This is a little transmitter that could be put into a plastic Easter egg. It delivers approximately 1 W of measured rf output into a 50-Ω dummy load, and creates no heating problems with the circuit. The crystal is a series fundamental type, and the power source provides 9 V with a 2-A supply. The transmitter can operate at another frequency, but C1 and C2 might have to be changed for the circuit to work properly. The secondary of L1 is wound over the center of a 10-μH coil, with five turns of #30 enameled wire.

LOW-COST HALF-DUPLEX INFORMATION TRANSMISSION LINK



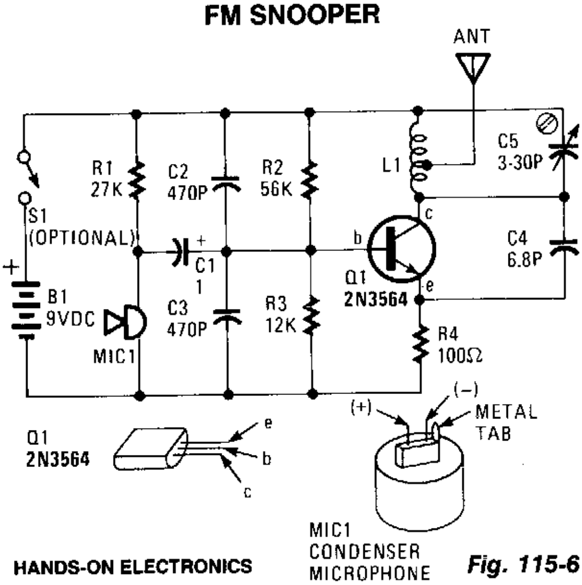
GE

Fig. 115-15

In a half-duplex system, information can flow in both directions, but only one direction at any given time. The conventional method of building a half-duplex link requires a separate emitter and detector, connected with directional couplers, at each end of the fiber. The GFOE1A series of infrared emitting diodes are highly efficient, long-lived emitters, which are also sensitive to the 940 nm infrared they produce. Biased as a photodiode, they exhibit a sensitivity of about 30 nA per μW irradiation at 940 nm. In a suitable bias and switching logic network, they form the basis for a half-duplex information link. A half-duplex link, illustrating the emitter-detector operation of the GFOE1A1, is shown.

This schematic represents a full, general purpose system, including: approximately 50-dB compliance range with 1-V rms output, passive receive, voice-activated switching logic, 100 Hz to 50 kHz frequency response, and inexpensive components and hardware. The system is simple, inexpensive, and can be upgraded to provide more capability through use of higher gain bandwidth amplifier stages. Conversely, performance and cost can be lowered simply by removing undesired features.

FM SNOOPER



HANDS-ON ELECTRONICS *Fig. 115-6*

The FM Snooper is an FM transmitter that radiates a continuous wave whose frequency is altered according to the sound waves striking the microphone. An ordinary FM broadcast receiver detects the transmitter's output carrier. When 9-V battery, B1, is connected, a brief surge of current flows from the collector to the emitter of Q1, causing an alternating current, shock oscillation in the resonant LC circuit, to flow back and forth between L1 and C5. So, by varying the value of C5, you can tune the oscillations to the exact frequency desired.

Although tuning capacitor C5 accounts for the major part of the tuning capacitance, the capacitance between the base and the collector of Q1 has a small, but noticeable, effect on the oscillation frequency. That capacitance, which is known as the *junction capacitance*, is not a fixed value, but instead varies when the voltage on the base of the transistor varies. Sound waves striking the microphone induce a voltage that varies in time with the sound. That voltage is applied via C1 to the base of Q1, thereby frequency modulating the transmitter.

VHF TONE TRANSMITTER

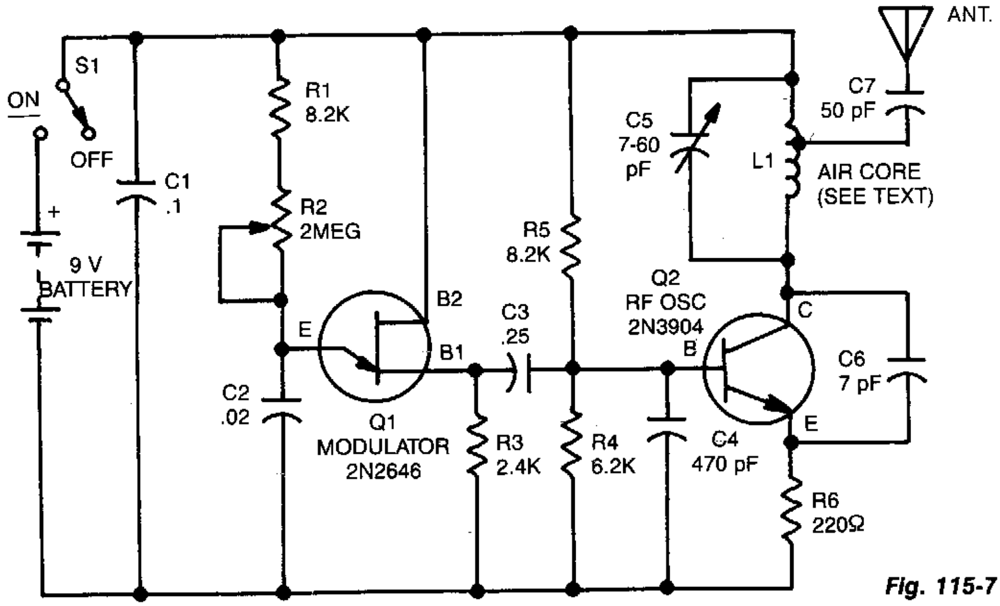


Fig. 115-7

The range of this transmitter is about 50 feet with a short whip antenna. The tone generator, is made up of a unijunction transistor, Q1, and R1, R2, R3, and C2. Transistor Q1 pulses on and off at a rate determined by the time constant of R1 and R2, together with the capacitance of C2 and the B1-emitter junction of Q1. Trimmer potentiometer R2 determines the frequency of the tone generated and allows a range of approximately 100 Hz to over 5 kHz.

Transistor Q2 is the rf oscillator. Its frequency is set by tuned circuits consisting of L1, C5, C6, and the interelectrode capacitance of Q2. The values shown will give a tuning range of about 55 to 108 MHz. Capacitor C6 provides positive feedback from the emitter to the collector of Q2, for oscillation. The audio tone generated by Q1 is applied to the base of Q2, causing the collector current to vary at the frequency of the tone, yielding an amplitude-modulated (AM) signal. This, in turn, varies Q2's collector-to-emitter capacitance, which makes up part of the tuned circuit, and causes the output frequency to vary similarly, producing a frequency-modulated (FM) signal, as well. The rf signal is coupled to the antenna through capacitor C7.

Coil L1 consists of five turns of #18 bare wire, close-wound on a piece of 1/4-inch wooden dowel. The length of the winding is about 1/4 inch. One end of capacitor C7 is soldered to the coil, one turn away from the 9-V supply end, and the other capacitor end is connected to the antenna. To adjust the vertical height and linearity of a TV set, place the tone transmitter near the set and use R2 to select the number of horizontal bars to be displayed. Once the picture is steady and the bars are sharp, adjust the set's vertical controls, so that all the bars are of the same height and are evenly spaced.

The fact that both AM and FM signals are generated makes it possible to use this circuit to check almost any receiver within the transmitter's frequency range. A TV set's sound section (discriminator) will reject the AM portion of the signal, but its video section will respond to it. Similarly, the TV sound section and FM receivers will respond to the FM signal produced.

LOW-FREQUENCY TRANSMITTER

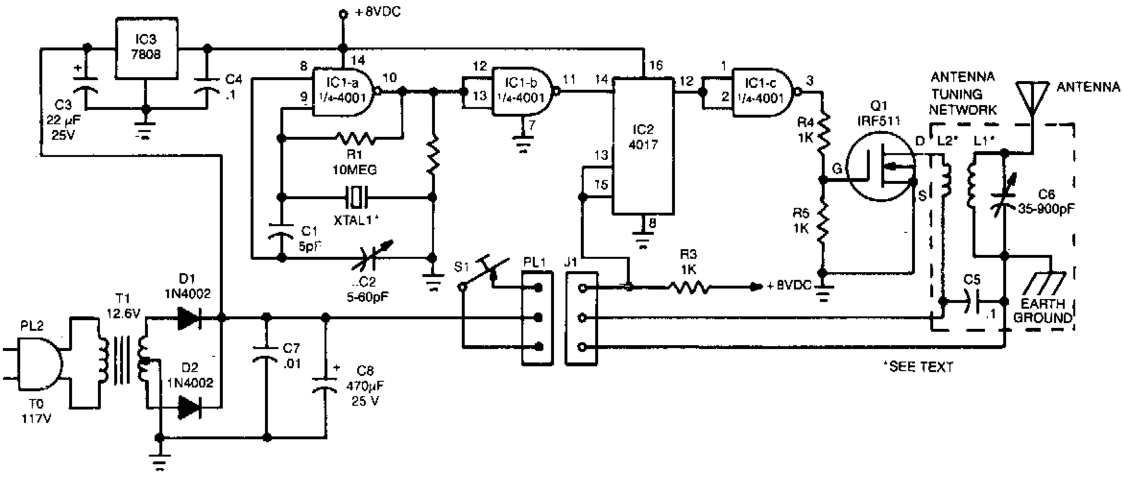
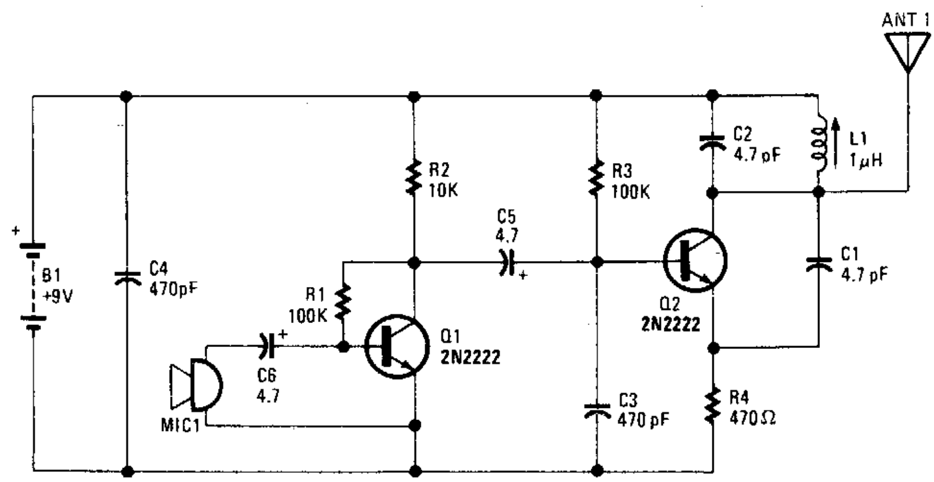


Fig. 115-8

The crystal oscillator, which uses two sections of IC1, a 4001 quad 2-input NOR gate, is a standard and reliable design. The oscillator's 1.85-MHz, square-wave output feeds IC2, a 4017 divide-by-10 counter. The count enable and reset terminals, pins 13 and 15, are normally held high by resistor R3, and the counter is activated by bringing those pins low by closing telegraph key S1—an arrangement that guarantees that the final state of IC2 pin 12 is always high. The high on IC2 pin 12 is inverted by a third section of the 4001, IC1c, to prevent dc current flow through power amplifier Q1 during key-up periods.

WIRELESS FM MICROPHONE

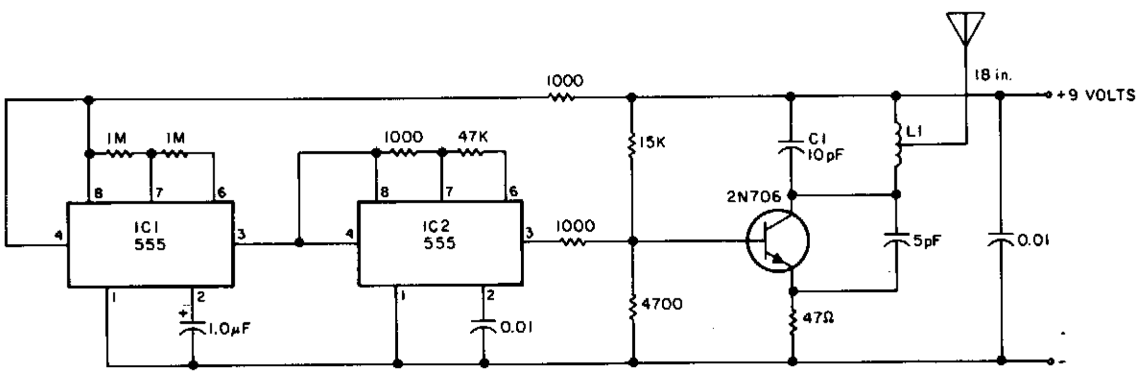


POPULAR ELECTRONICS

Fig. 115-9

Transistor Q1 acts as an amplifier for condenser microphone MIC1. The output of Q1 is applied to the base of transistor Q2 through a 4.7-μF capacitor. C2 and L1 form an LC tank circuit, which is used to set the frequency at which the transmitter operates. Coil L1 is a variable inductor, centered a bit below 1 μH, that is used to adjust the modulating frequency of the circuit. Capacitors C1 and C2 are 4.7 pF units. A lower value can be used to raise the circuit's operating frequency. The microphone and Q1 provide a varying voltage at the base of Q2, with the output of Q2 applied to the LC tank circuit. That causes a modulating action in the tank circuit that, when applied to the antenna, a short piece of wire 6- to 8-inches long, will provide a good, clear FM signal somewhere in the range of 88 to 95 MHz with a range of about 100 feet.

BEACON TRANSMITTER

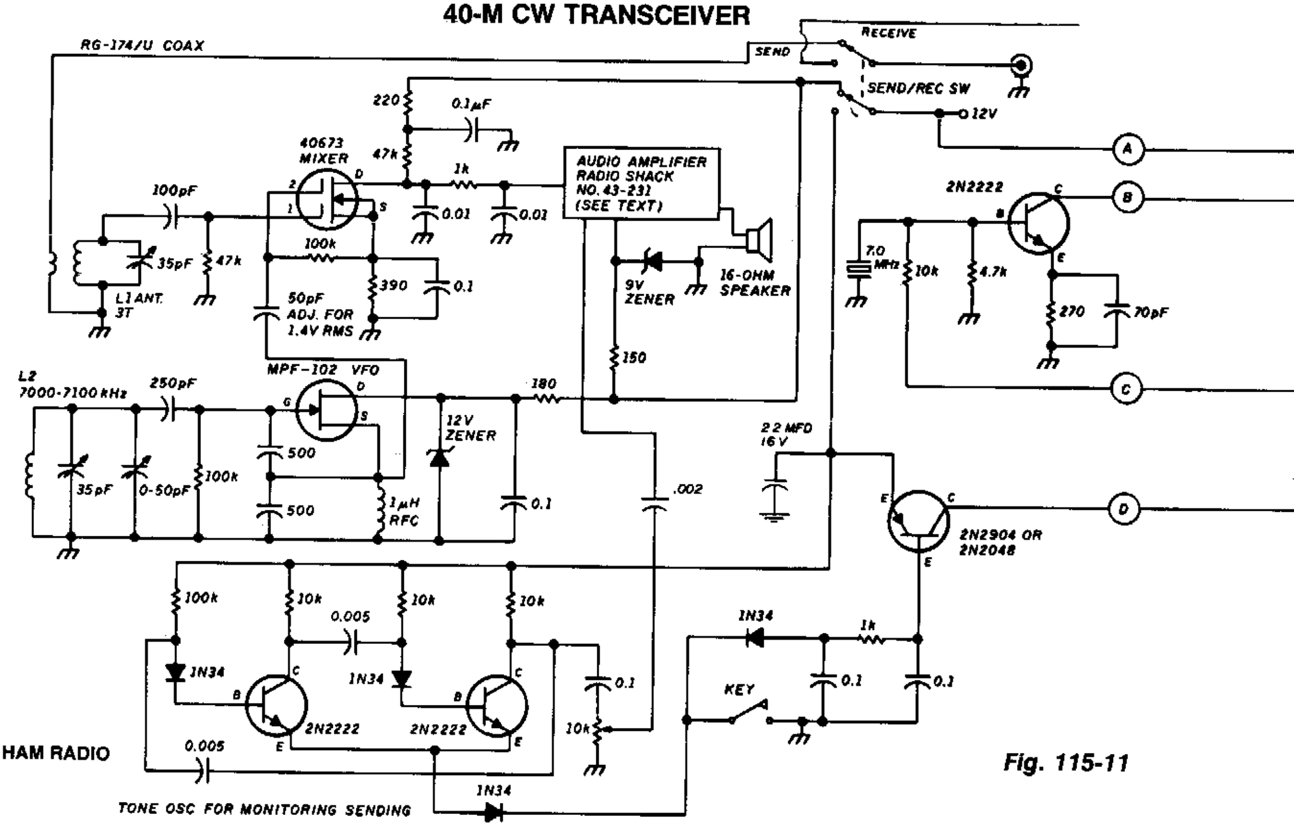


73 AMATEUR RADIO

Fig. 115-10

This transmitter can be used for transmitter hunts, for remote key finding, or for radio telemetry in model rockets. It can be tuned to the two meter band or other VHF bands by charging C1 and L1. L1 is four turns of #20 enameled wire airwound, 0.25 inch in diameter (use a drill bit), 0.2 inch long, center-tapped. The antenna can be 18 inches of any type of wire. IC2 functions as an audio oscillator that is turned on and off by IC1 about once per second. The range of the transmitter is several hundred yards.

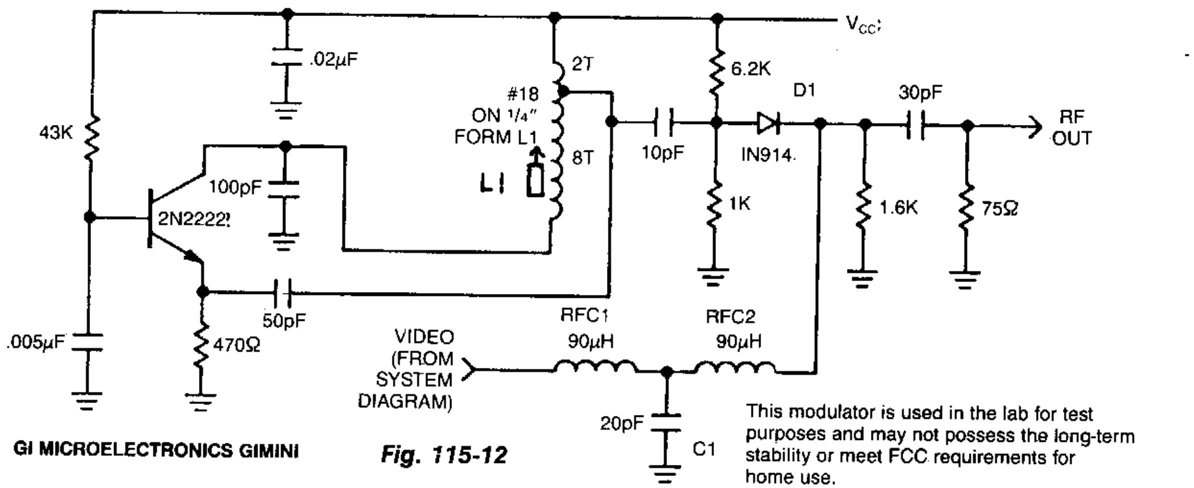
40-M CW TRANSCEIVER



HAM RADIO

Fig. 115-11

VHF MODULATOR

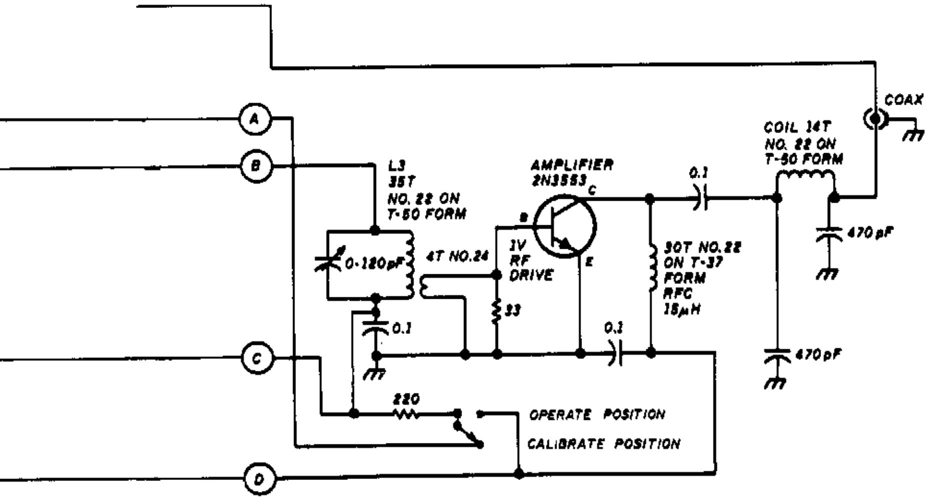


GI MICROELECTRONICS GIMINI

Fig. 115-12

This modulator is used in the lab for test purposes and may not possess the long-term stability or meet FCC requirements for home use.

This circuit uses an oscillator (2N2222) and a diode D1 as a nonlinear mixer. The frequency is set by a slug in L1. RFC1, C1, and RFC2 form a low-pass filter to pass video and block rf from the video source.



The unit consists of a direct conversion receiver and 1-W transmitter. The direct conversion receiver VFO is tuned just off frequency from the incoming signal. This difference in frequency produces a clean, strong, and solid audio tone signal. Detect the resonant frequency of the transmitter VFO by using the GDO as a field-strength meter.

Because of the large capacitance in the Colpitts VFO, the tuning coil will have fewer turns than the mixer coil. Use the capacitance shown for the VFO gate to ground and to the coil. It will effect the frequency and output. You'll need 1.4-V rms on pin 2 of the mixer to get a good signal from the VFO. The 1000-Ω resistor and 0.01-μF capacitors act as an rf filter from the mixer output.

WIRELESS FM MICROPHONE

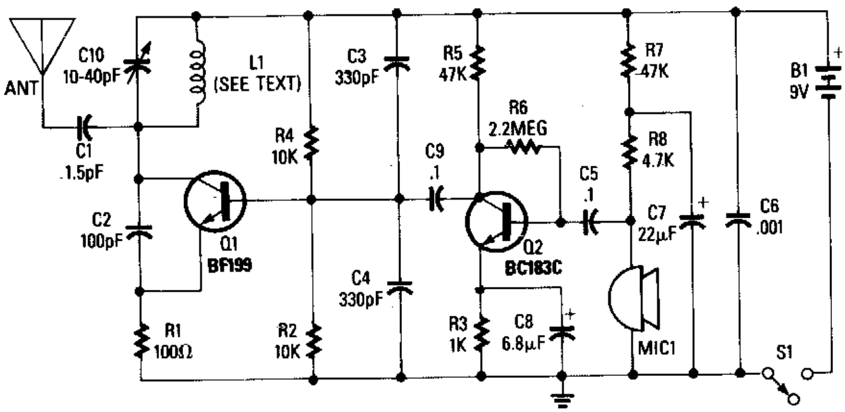


 Fig. 115-13

Adjustable capacitor C10, and coil L1 form a tank circuit that, in combination with Q1, C2, and R1, oscillates at a frequency on the FM band. The center frequency is set by adjusting C10. An electret microphone, M1, picks up an audio signal that is amplified by transistor Q2. The audio signal is coupled via C9 to Q1, which frequency modulates the tank circuit. The signal is then radiated from the antenna. The circuit can operate from 9 – 12 Vdc.

902-MHz CW TRANSMITTER

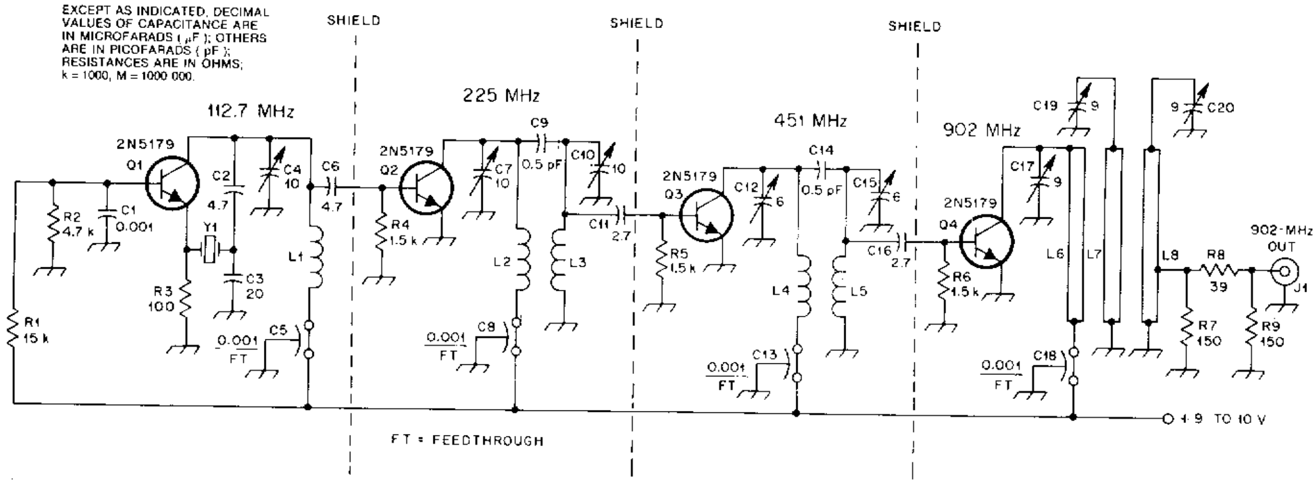


Fig 2—Schematic diagram of the exciter. Resistors are ¼-W carbon composition. Capacitors are 50-V epoxy-coated ceramic types unless otherwise noted.

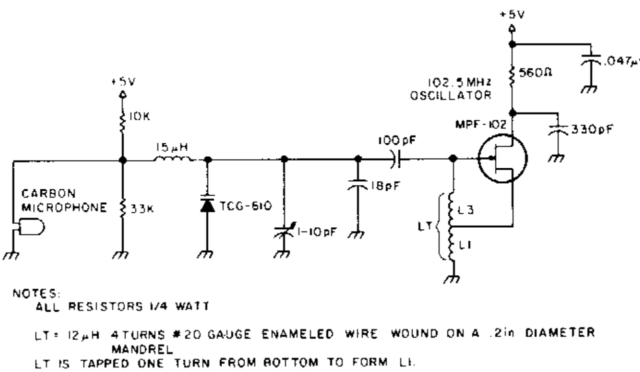
C4, C7, C10—1.5-15 pF miniature air-variable capacitor (Trim-tronics 10-1120-25015-000 or equiv).
C5, C8, C13, C18—470- to 1000-pF ceramic feedthrough capacitor, solder-in type preferred.
C9, C14—0.5-pF "gimmick" capacitor (see text).
C12, C15—1-6 pF miniature air-variable capacitor (Trim-tronics 10-1120-25006-000 or equiv).
C17, C19, C20—0.6-9 pF ceramic piston trimmer capacitor (Voltronics EQT9 or equiv).
J1—Chassis-mount female BNC connector (UG-1094 or equiv).
L1— 5t no. 22 tinned wire, 0.228-in ID (no. 1 drill), spaced 1 wire dia.
L2, L3—4t no. 18 tinned wire, ¼-in ID, spaced 1 wire dia.
L4, L5—2t no. 18 tinned wire, ¼-in ID, spaced 1 wire dia.
L6, L7, L8—Inductor made from copper strap, 1-in long × 1/8-in wide. See text and Fig 3 for details.
Q1-Q4—2N5179 transistor.
Y1—Fifth-overtone crystal, 80.545 MHz, or seventh-overtone crystal, 112.763 MHz, HC-25 holder, series resonant, 0.005% (avail from JAN Crystals, 2400 Crystal Dr, Ft Meyers, FL 33906 tel 800-237-3063).

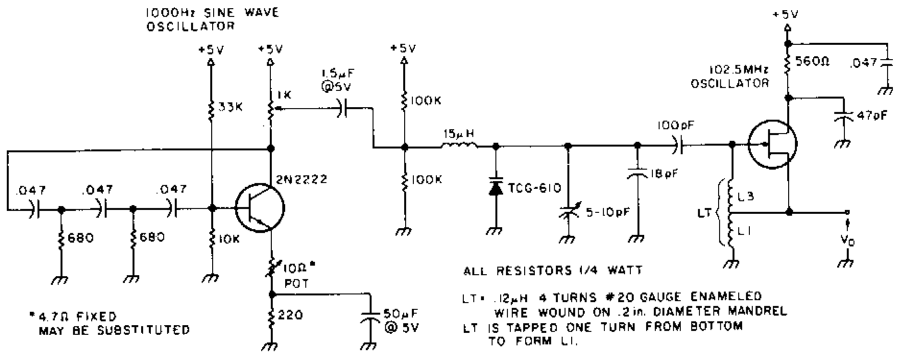
QST

Fig. 115-14

The oscillator, Q1, is a standard overtone circuit. A fifth-overtone crystal, 80.545 MHz, is operated on the seventh overtone, 112.763 MHz. C6 couples the output of the oscillator to Q2, which operates as a doubler to 225.5 MHz. A double-tuned circuit using C7, L2, L3, C10 is used in the collector of Q2 to reduce the level of the 112-MHz oscillator signal. The output of Q2 is capacitively coupled at C11 to the base of Q3. The double-tuned circuit in the collector of Q3 with C12, L4, L5, C15, is tuned to 451 MHz. A small capacitance, 2.7 pF, couples the 451-MHz signal to the base of another 2N5179, Q4, which doubles the signal to 902 MHz. The output of the 902-MHz doubler has a triple-tuned circuit using C17, L6, C19, L7, C20, L8 in its collector.

A ONE-TRANSISTOR FM TRANSMITTER



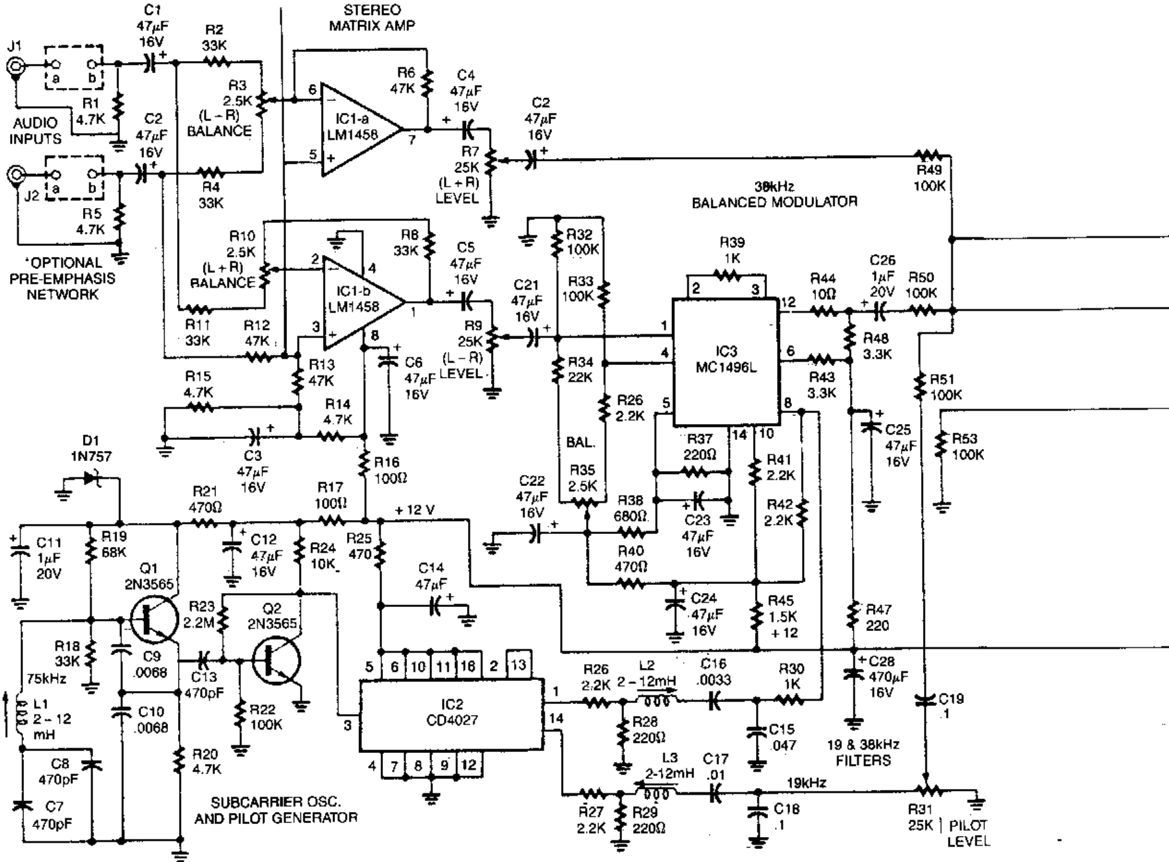


73 AMATEUR RADIO *Fig. 115-15*

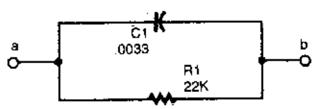
The 2N2222 circuitry is a three-element, phase-shift oscillator circuit, designed to yield a 1,000-Hz sine wave. The 1,000-Hz sine wave is then applied to the TCG-610 varactor diode, 6 pF at 4 V, which changes the tank capacitance, thus varying the rf oscillator frequency at a 1,000-Hz rate. The 1,000-Ω potentiometer in the collector circuit can be adjusted to enable the desired frequency modulation level.

The Hartley rf oscillator, designed around a readily available MPF-102 JFET, has an output that should be relatively stable if it is enclosed in a metal box, thus minimizing changes in tank capacitance. The completed transmitter has a range of 30 feet when not enclosed—without an antenna.

FM MULTIPLEX TRANSMITTER



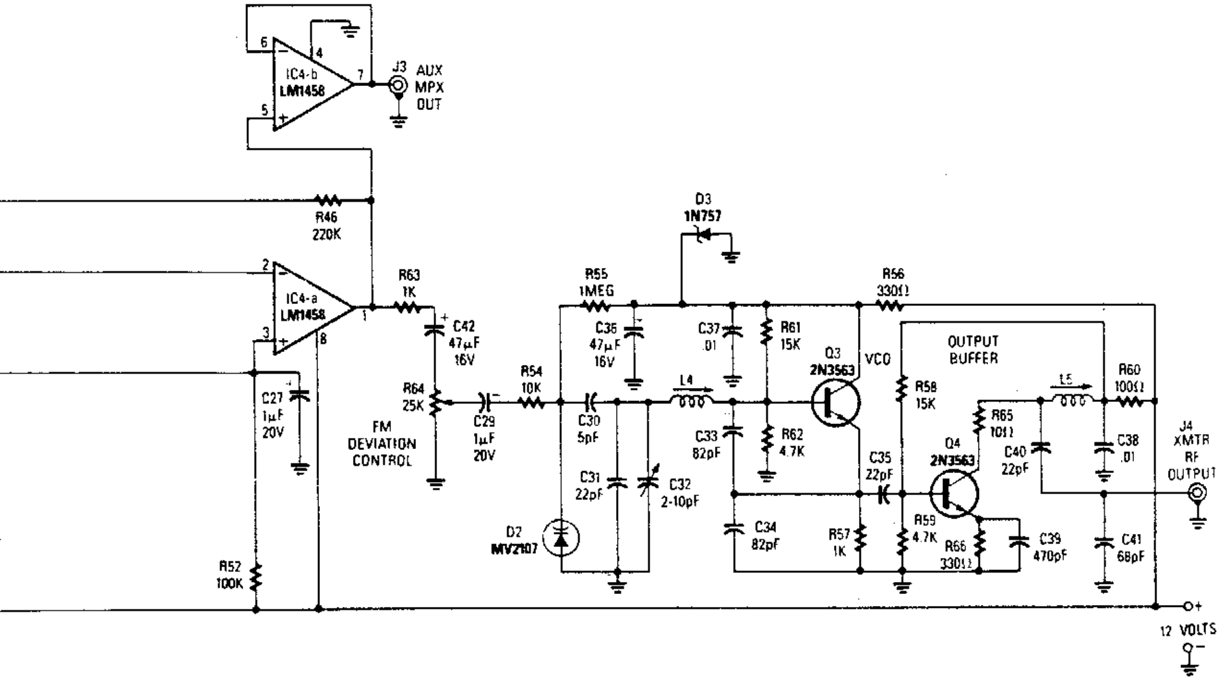
THIS STEREO FM TRANSMITTER is capable of transmitting a stereo signal up to a hundred feet.

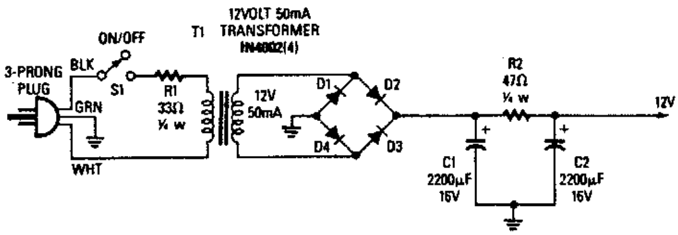


PRE-EMPHASIS NETWORK (OPTIONAL) NOT ON PCB

THIS PRE-EMPHASIS NETWORK can be added to the audio inputs of the MPX transmitter, if necessary.

RADIO-ELECTRONICS *Fig. 115-16*

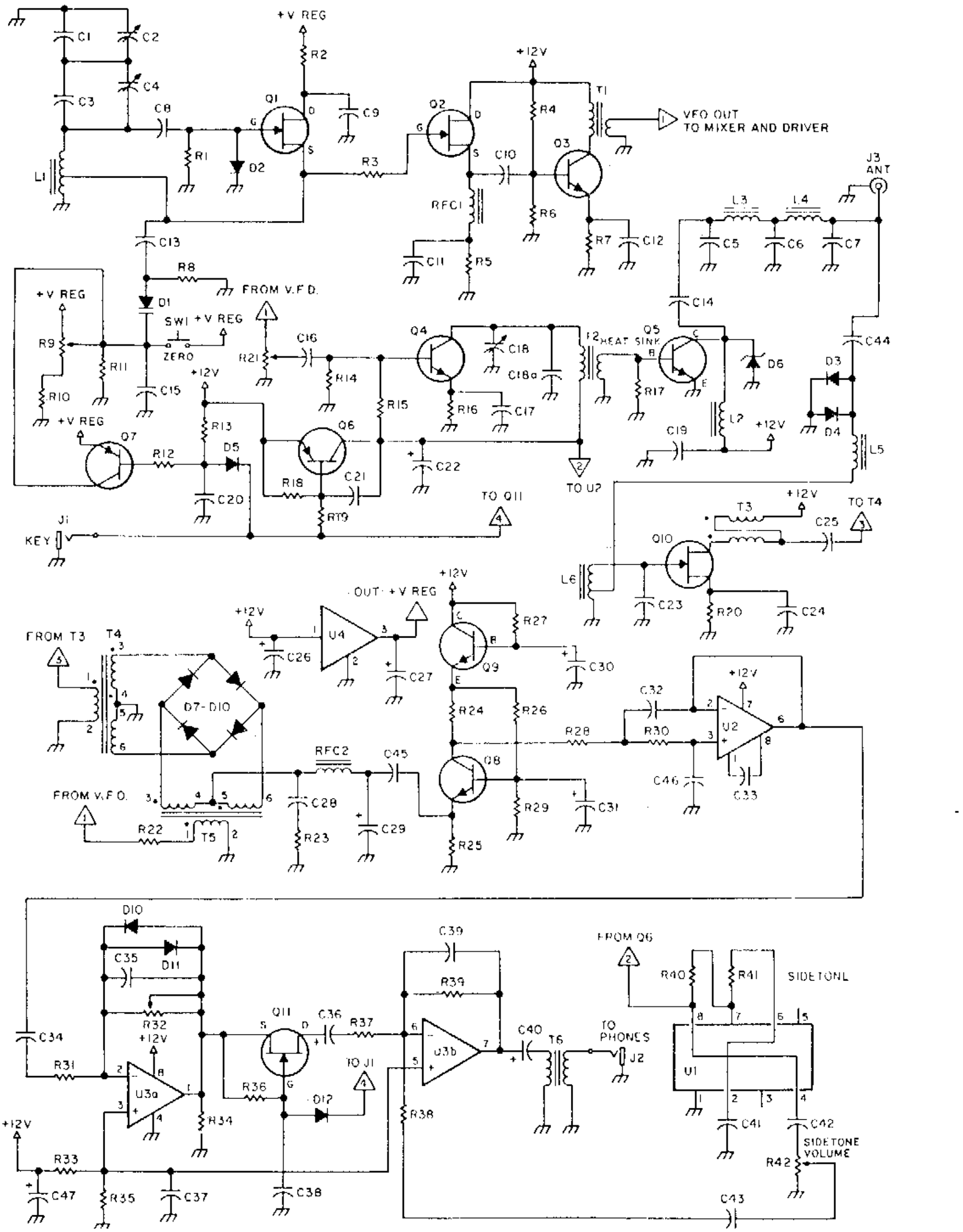




THIS POWER SUPPLY can be used if you do not want to power the transmitter with batteries.

This transmitter has a range of up to 100 feet. It generates a complete multiplex stereo signal and is useful for cordless headphone applications in which an inexpensive socket stereo receiver can be used. It can also be used as an FM multiplex generator for receiver alignments. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

QRP CW TRANSCEIVER



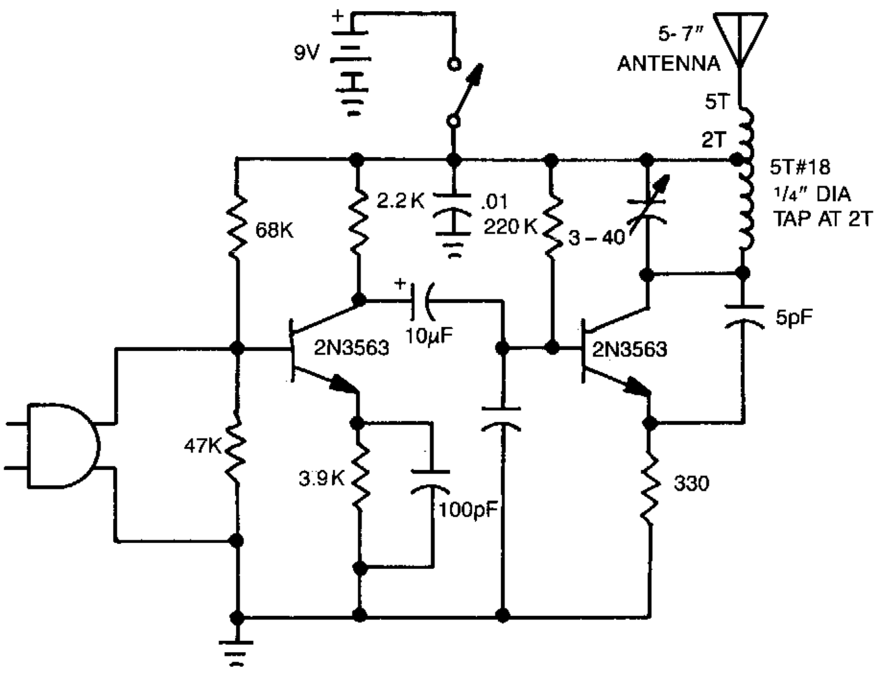
73 AMATEUR RADIO

Fig. 115-17

This is a 3-W, single-circuit board, VFO-controlled CW transceiver for 40 or 30 meters, featuring a direct-conversion receiver with audio filtering, Receiver Incremental Tuning (RIT), and speaker level audio volume. The transmit frequency is generated by Q1 and its associated components in the VFO. The buffer, Q2, isolates the oscillator from the other circuitry to help keep the VFO stable. Q3 builds up the signal to a more usable level. The driver, Q4, amplifies the signal. The final, Q5, amplifies it to the 3-W level.

Key the transmitter by turning the power to the driver on and off, using Q6 as a switching transistor. Select the frequency by varying the tuning capacitor, C2. The VFO frequency feeds into the diode-ring mixer, and is mixed with the incoming 7- or 10-MHz signal. The *difference*, or *produce*, is the audio frequency. The post-mixer circuitry amplifies the audio signal to speaker level: Q8 preamplifies the signal a little, U2 is an audio filter that attenuates the audio signals above about 700 Hz, and U3 amplifies the signal from the audio filter to listening level.

WIRELESS FM MICROPHONE



Reprinted with permission of William Sheets.

Fig. 115-18

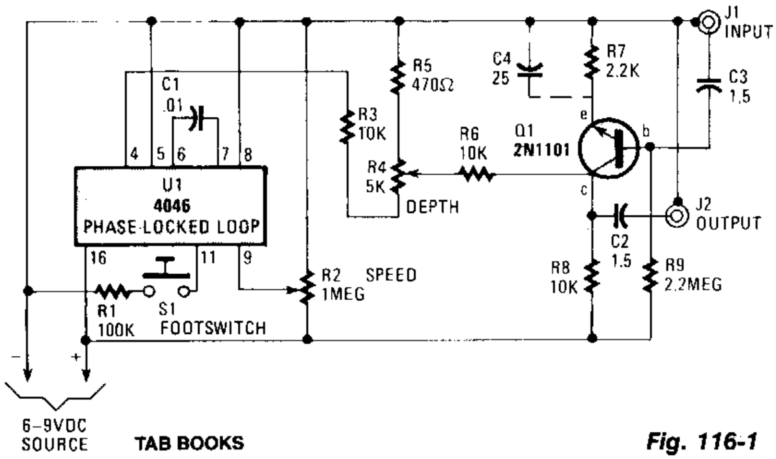
Use standard rf wiring precautions. The best speech clarity is obtained by using an electret microphone. For music reproduction, substitute a dynamic mike element.

116

Tremolo Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

TREMOLO

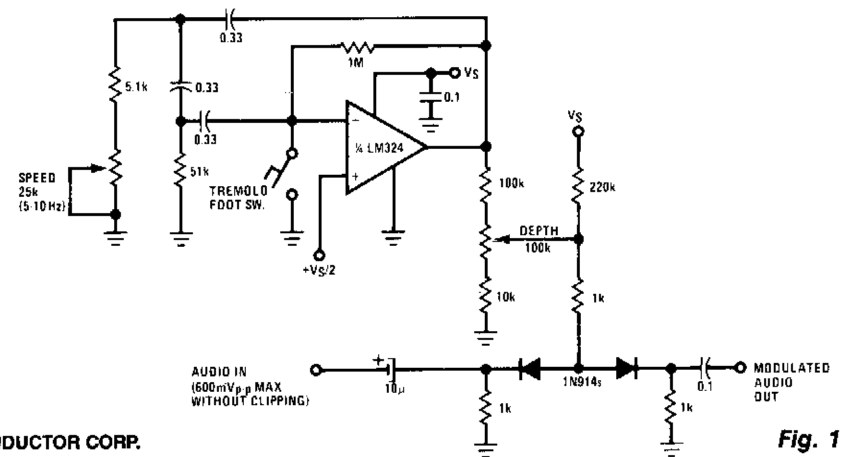


TAB BOOKS

Fig. 116-1

The VCO of U1, a phased-locked loop, produces a low-frequency square-wave signal from 5 Hz to 2500 Hz, which is controlled by varying the voltage to pin 9 via R2. The frequency threshold is set by R1 and C1. Increasing the value of R1 increases the frequency, and decreasing the value of C1 decreases the threshold frequency. Transistor Q1 is operated as an amplifier with the Q point at 4 V, using a transistor with a voltage gain (beta) of 100. The gain without tremolo is about 2 V. With bypass capacitor C4 in the circuit, the gain is set at 33 dB. The instrument drive cannot be greater than 30 mV with C4 in the circuit (no tremolo), to avoid distortion. Without C4, the drive can be 1.5 V maximum with no distortion. U1's output is coupled to the collector of transistor Q1 via depth control R4. The square-wave signal pulls the audio at a frequency determined by R2. The oscillator is activated by foot switch S1.

TREMOLO

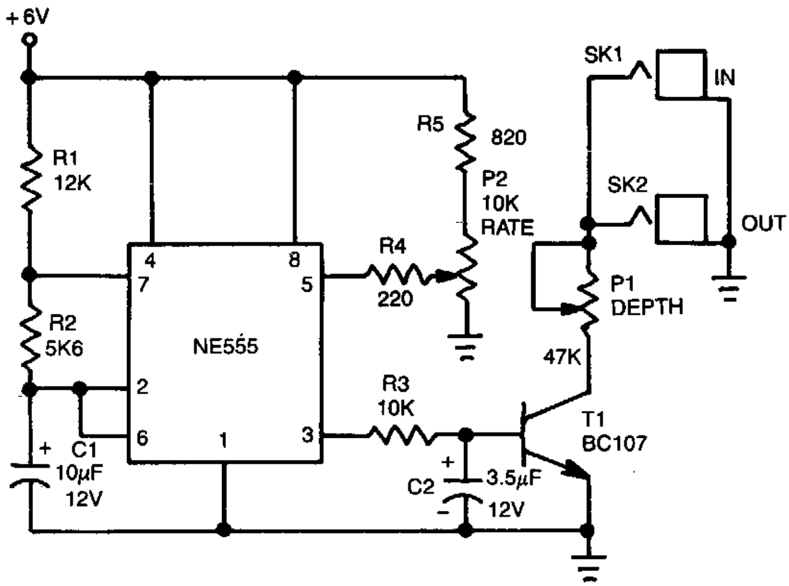


NATIONAL SEMICONDUCTOR CORP.

Fig. 116-2

A phase shift oscillator using the LM324 operates at an adjustable rate, 5–10 Hz, set by the speed pot. A portion of the oscillator output is taken from the depth pot and used to modulate the on resistance of two 1N914 diodes operating as voltage controlled attenuators. Care must be taken to restrict the incoming signal level to less than 0.6 V pk-pk, or undesirable clipping will occur.

ELECTRONIC TREMOLO



ELECTRONIC ENGINEERING

Fig. 116-3

The tremolo effect is generated by a repeating volume change at a rate usually between 1 and 15 Hz. The timer produces a low frequency square wave that is smoothed by a simple rc integrator. This varying signal modulates the signal input from the instrument. Transistor T1 is used as a voltage controlled resistor. The output of the circuit is connected in parallel to the output of the instrument. Potentiometer P1 provides depth control by adjusting the amplitude of the modulating waveform applied to the instrument. The rate control frequency is set by potentiometer P2.

TREMOLO

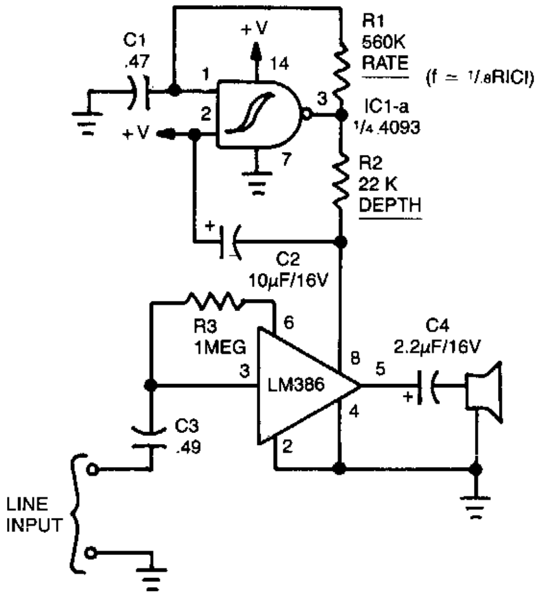


Fig. 116-4

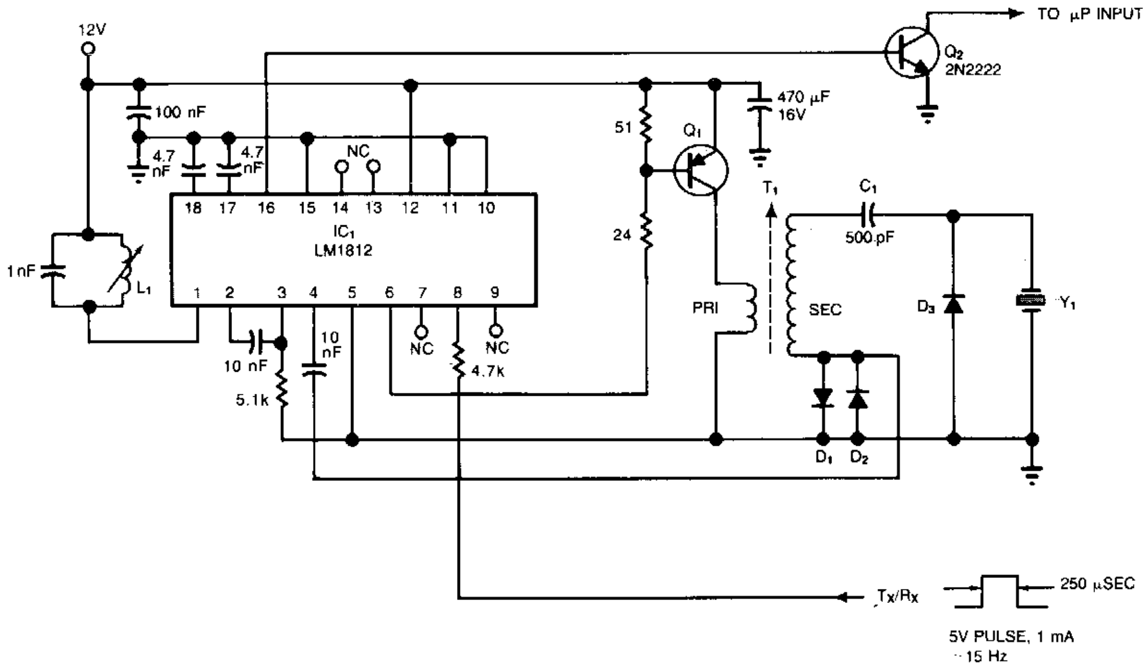
This simple circuit can color the sound coming from your audio system. Clocking for the circuit is provided by an oscillator built from one quarter of a 4093 quad NAND Schmitt trigger. With the component values shown, it will run at about 5 Hz. The clock frequency is fed to the gain control, pin 8, of an LM386 amplifier. Tremolo is produced by varying the amplifier gain. A trimmer potentiometer can be put in series with R1, to easily experiment with different rates. To experiment, make R1 about 100-KΩ and use a 1-MΩ trimmer. That allows frequencies from about 2 to 20 Hz to pass. Resistor R2 is the depth control. It controls the degree of tremolo. To adjust, put a trimmer in series with R2. Make R2 a 5-KΩ unit and use a 50-KΩ trimmer. Since the tremolo clock uses the gain-control pin of the amplifier, change the value of capacitor C4 in order to change the gain of the amplifier. Make C4 larger to increase the gain or smaller to decrease it. But, don't go any lower than 0.1 μF because you'll be cutting into the bottom-end frequency response.

117

Ultrasonics

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

RANGING SYSTEM



PART	PARTS LIST DESCRIPTION	MANUFACTURER
L_1	15.8 mH ADJUSTABLE #CLN-2A900HM	TOKO AMERICA (SKOKIE, IL)
T_1	PRIMARY 8 TURNS #24* SECONDARY 110 TURNS #30* POTCORE RM8P-A630-3B7 BOBBIN RM8 PCB1-4 CLIPS 991-393-00	FERROXCUBE (SAUGERTIES, NY)
X_1	TRANSDUCER	POLAROID (CAMBRIDGE, MA)
Q_1	D45C6 I_C = 5A MIN; V_{CBO} = 40V MIN	NATIONAL SEMICONDUCTOR (SANTA CLARA, CA)

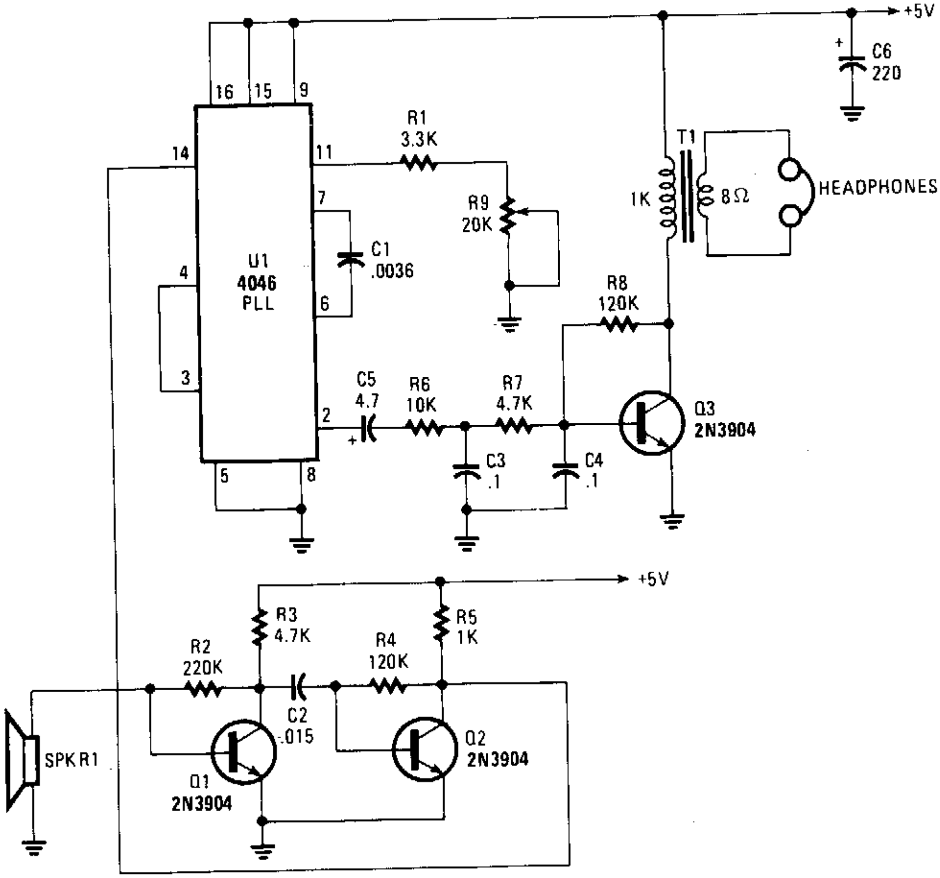
*IF MACHINE WOUND, SLIGHTLY LARGER WIRE SIZES MAY BE USED.

EDN

Fig. 117-1

Combine an electrostatic transducer with an ultrasonic transceiver IC to build a ranging system that senses objects at distances from 4 inches to more than 30 feet. Transducer Y1's broadband characteristic simplifies tuning. The secondary of T1 resonates with the 500-pF capacitor C1 at a frequency between 50 and 60 kHz. You tune L1 to this frequency by using an oscilloscope to note the maximum echo sensitivity at pin 1. Step-up transformer T1 provides 150-V bias for the transducer.

ULTRASONIC RECEIVER



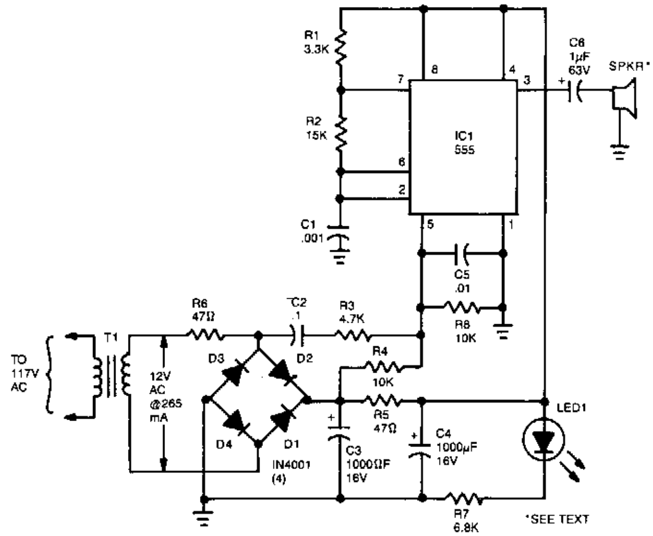
POPULAR ELECTRONICS

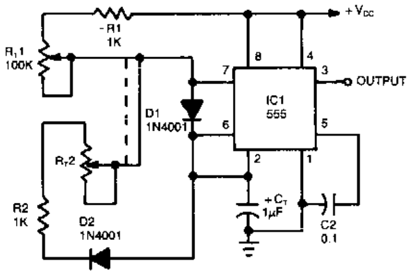
Fig. 117-2

The 4046 PPL is used as the heart of a tunable ultrasonic receiver that can be used to locate unheard ultrasonic sounds. The receiver might also be used, along with a simple ultrasonic generator, to send and receive Morse code. The incoming ultrasonic signal is picked up by piezo speaker SPKR1, and amplified by transistors Q1 and Q2. The output is fed to the phase comparator input of U1 at pin 14. The chip's interval VCO is tuned by turning potentiometer R9.

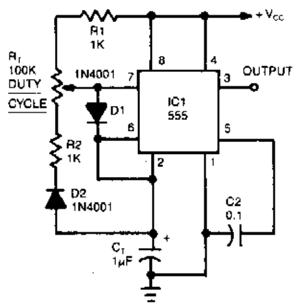
If a 20-kHz signal is picked up by SPKR1 and the VCO is tuned to produce a 19-kHz signal the difference output at pin 2 will be 1 kHz. That 1-kHz signal is amplified by Q3 and coupled through T1 to a pair of headphones. If the received frequency increases to 22 kHz, a 3-kHz tone is heard in the headphones. With the values given in the parts list for C1, R1, and R9, the VCO can be tuned from 12 to well over 42 kHz, which should cover just about anything the piezo sensor can respond to.

ULTRASONIC PEST-REPELLER





If you wish to vary the duty cycle while keeping the frequency constant, replace the timing resistors with a single potentiometer



If you wish to vary the frequency, but keep the duty constant, at about 50%, the 555 circuit can be modified as shown.

Fig. 117-3

This circuit is a 555 timer IC connected as a square-wave generator. Its base frequency is approximately 45 kHz, as determined by the values of R1, R2, and C1. The 45-kHz *carrier* is frequency modulated by a modified trapizoidal voltage waveform applied to pin 5 of the 555 timer. That modulating voltage is developed by a network consisting of C2, R3, and R4 connected across one leg of the bridge rectifier. The sweep is approximately 20 kHz on each side of the base frequency. The speaker is a 2-inch piezoelectric tweeter.

20-kHz ARC WELDING INVERTER

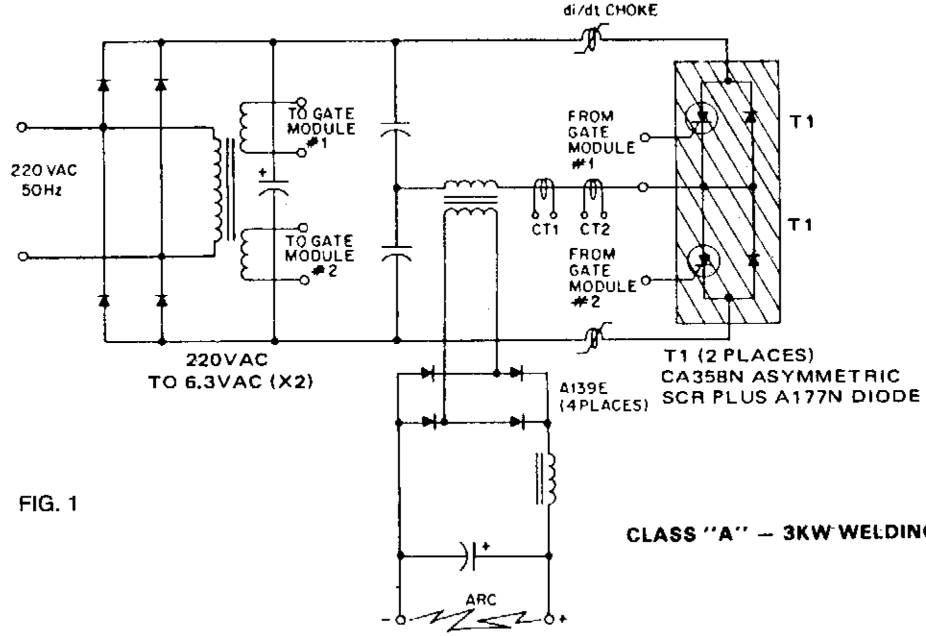


FIG. 1

CLASS "A" — 3KW WELDING INVERTER

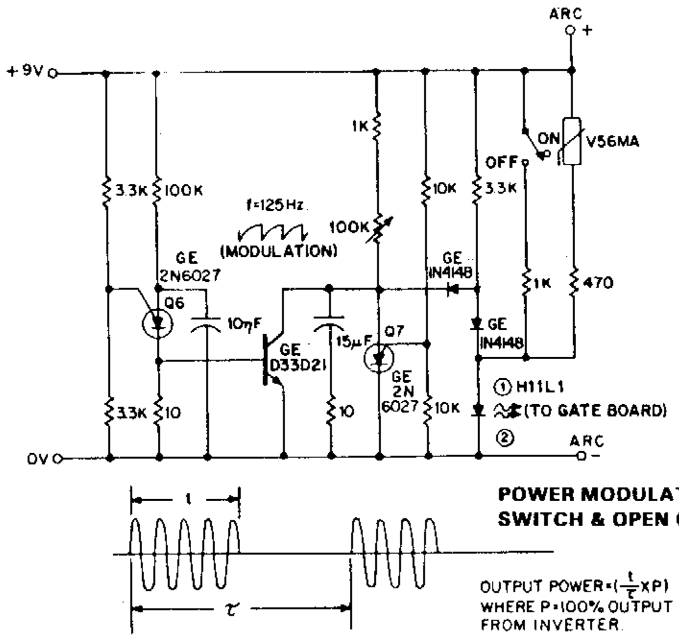


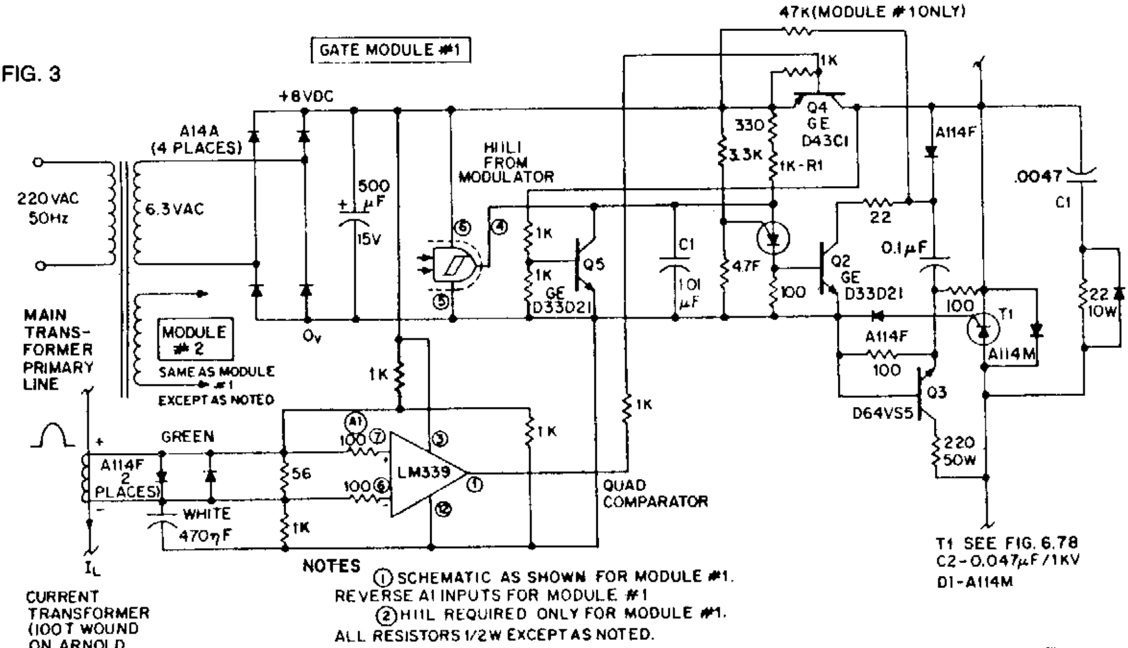
FIG. 2

POWER MODULATOR (WITH ON-OFF SWITCH & OPEN CIRCUIT PROTECTION)

OUTPUT POWER $=(\frac{t}{\tau} \times P)$
WHERE P=100% OUTPUT FROM INVERTER

The Class A series resonant inverter portrayed is well-known and respected for its high efficiency, low cost, and small size, provided that operating frequency is greater than about 3 kHz. The disadvantages are, at least in high power versions, the difficulty in effecting smooth RFI-free output voltage modulation

FIG. 3



GE

20KHz INVERTER GATE DRIVE MODULE ***Fig. 117-4***

without significant added complexity, and a natural tendency to *run away* under no-load (high Q) conditions.

The 20-kHz control circuit (see Fig. 2) overcomes these shortcomings by feeding back into the asymmetrical thyristor trigger pulse generators (see Fig. 3) signals that simultaneously shut the inverter down, when its output voltage exceeds a preset threshold, then time-ratio modulates the output. This feedback is accomplished with full galvanic isolation between input and output thanks to an H11L opto-Schmitt coupler. The fundamental 20-kHz gate firing pulses are generated by a PUT relaxation oscillator Q1. The pulses are then amplified by transistors Q2 and Q3. The 20-kHz sinusoidal load current flowing in the primary of the output transformer is then detected by current transformer CT1, with op amp A1 converting the sine wave into a square wave, whose transitions coincide with the load current zero points.

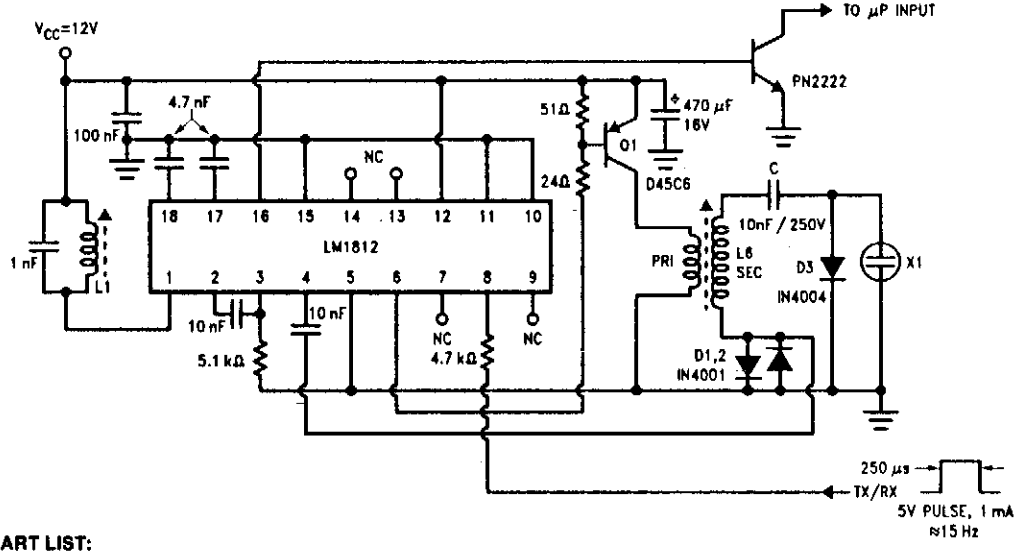
Consequently, each time the output current changes, phase A1 also changes state and, via transistor Q4, either connects the thyristor gate to a −8 Vdc supply for minimum *gate assisted* turn-off time and highest reapplied dV/dt capability or disables this supply to prepare the thyristor for subsequent firing.

Modulation intelligence is coupled into this same H11L through two additional PUTs, Q6 and Q7, Q6 oscillates at a fixed 1.25 kHz, which establishes the modulation frequency. The duty cycle is determined by a second oscillator, Q7, whose conduction state, on or off, establishes or removes current from the H11L diode. With a fundamental inverter frequency of 20 kHz and a modulation frequency of 1.25 kHz, the resultant time ratio-controlled power output is given by:

$$P_{OUT} = \left(P_M \times \frac{t}{T} \right)$$

where P_M = 100% continuous output power. Minimum power is one cycle of 20 kHz (50 μs) in the 1.25-kHz modulation frame (800 μs), that is, 6.25% P_M.

ULTRASONIC TRANSCEIVER



PART LIST:

L1	15.8 mH adjustable. #CLN-2A900HM		TOKO
L6	PRIMARY	8 TURNS #24*	
	SECONDARY	110 TURNS #30*	
	POTCORE	RM8P-A630-3B7	FERROXCUBE
	BOBBIN	RM8 PCB1-4	
	CLIPS	991-393-00	
X1	POLAROID TRANSDUCER		POLAROID
Q1	D45C6 $I_C = 5A_{min}$ $V_{CBO} = 40V_{min}$		NSC
C	250V, 10 nF MYLAR		—

*If machine wound, slightly larger wire sizes may be used.

Toko America, Inc. 1250 Feehanville Drive
Mount Prospect, IL 60056
Tel. (312) 297-0070

Ferroxcube 5083 Kings Highway,
Saugerties, NY 12477
Tel. (914) 246-2811

Polaroid Corp. Commercial Battery Division
784 Memorial Drive,
Cambridge, MA 02139
Tel. (617) 577-2024

National Semiconductor Corp. 2900 Semiconductor Drive
Santa Clara, CA 95051
Tel. (408) 721-5000

NATIONAL SEMICONDUCTOR CORP. ***Fig. 117-5***

The LM1812 is a complete ultrasonic transceiver on a chip designed for use in a variety of pulse-echo ranging applications. The chip operates by transmitting a burst of oscillations with a transducer, then using the same transducer to listen for a return echo. If an echo of sufficient amplitude is received, the LM1812 detector puts out a pulse of approximately the same width as the original burst. The closer the reflecting object, the earlier the return echo. Echos could be received immediately after the initial burst was transmitted, except for the fact that the transducer *rings*.

When transmitting, the transducer is excited with several hundred volts peak to peak, and it operates in a *loudspeaker* mode. Then, when the LM1812 stops transmitting and begins to receive, the transducer continues to vibrate or ring, even though excitation has stopped. The transducer acts as a microphone and produces an ac signal initially the same amplitude as the transmit pulse. This signal dies away as is governed by the transducer's damping factor, but as long as detectable ringing remains, the LM1812's detector will be held on, masking any return echos.

SONAR TRANSDUCER/SWITCH

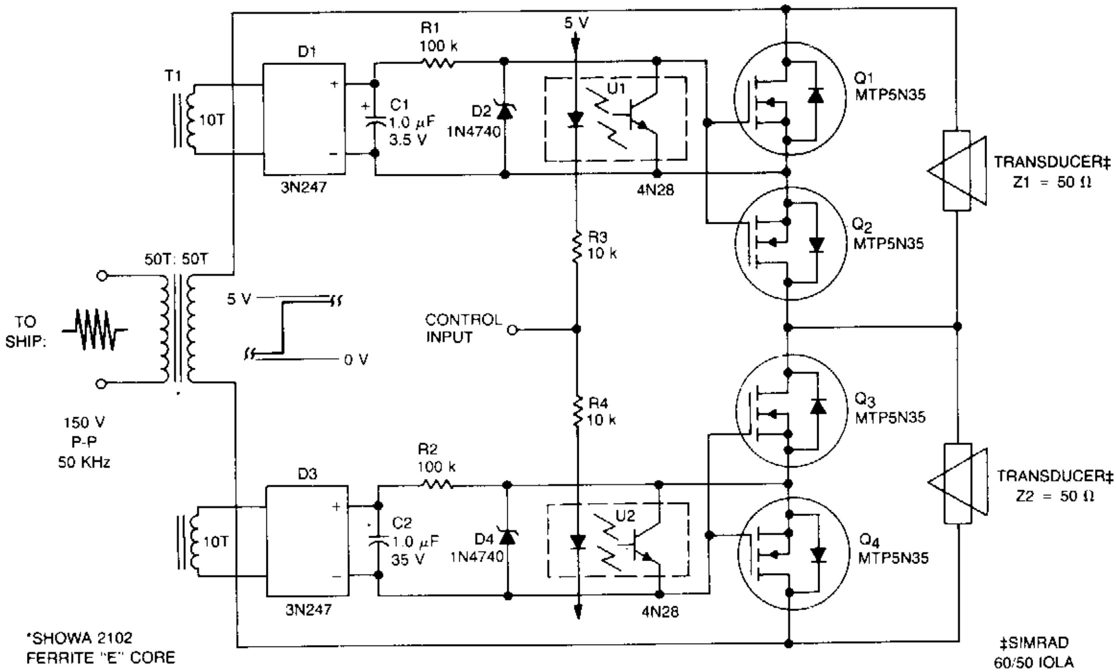


Fig. 117-6

This submersible sonar positioning apparatus generally consists of dual-opposed ultrasonic transducers, alternately excited, with return signals processed and displayed for observation and measurement. Typical transmitter frequencies range from 50 to 200 kHz and pulse widths can be varied from 0.3 to 5 ms, depending on depth and resolution requirements.

The input to the transducer/switch is transformer T1 which provides isolation and impedance matching. The turn ratio of the secondary windings depends on the peak-to-peak amplitude of the transmitter output into the specified load. The transmitted pulse that appears on the secondary winding charges capacitors C1 and C2 through bridge rectifiers D1 and D3. Zener diodes D2 and D4 limit the TMOS gate bias to 12 V; R1 and R2 limit the discharge current from C1 and C2.

The square-wave control input is applied to opto-isolators U1 and U2 through resistors R3 and R4. If the control input is 0 V, U1 is activated; when it changes to +5 V, U2 is activated. When U1 is activated, it saturates and reduces the gate bias to zero, turning Q1 and Q2 off. Q3 and Q4 remain on, effectively shunting transducer Z2. When U2 is activated, it saturates and reduces the bias to zero, turning Q3 and Q4 off. Q1 and Q2 remain on, effectively shunting transducer Z1.

120-kHz 500-W INDUCTION HEATER

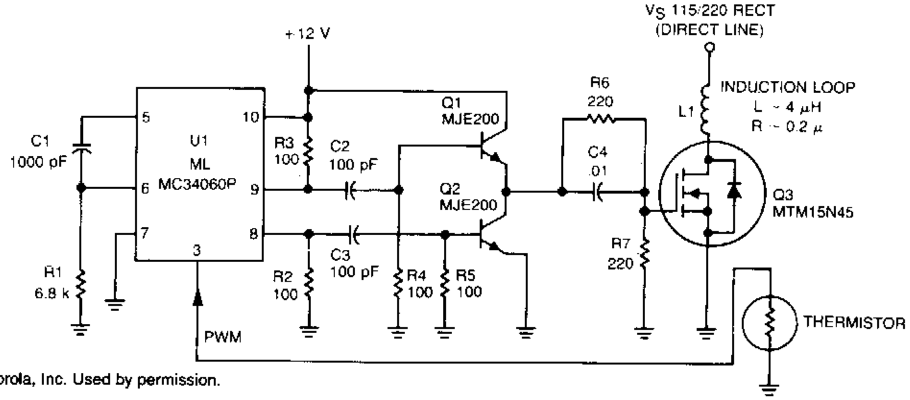
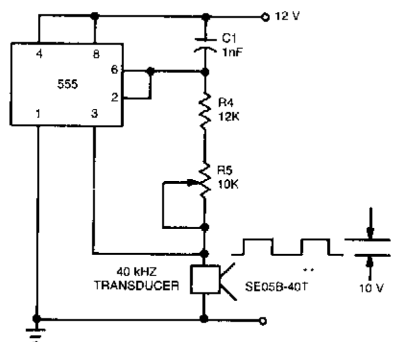


Fig. 117-7

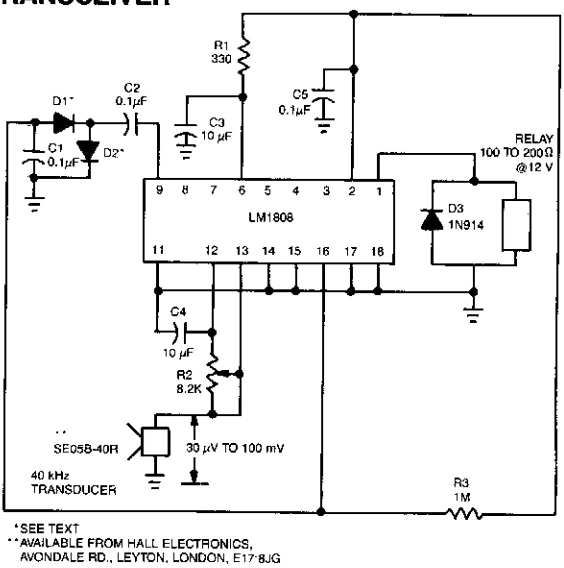
Variable width pulses with fast rise times are provided by U1, and MC34060 operating at 120 kHz, the optimum frequency for heating aluminum alloy containers. The pulse width is modulated by sensing the temperature of the target with a thermistor, using its negative temperature coefficient to change pulse duration. The MC34060 produces output pulses that are ac-coupled to push-pull MJE200 transistors Q1 and Q2. This IC provides the current needed to ensure fast switching for MTM15N45 TMOS power FET Q3.

The estimated efficiency is 80%, based on switching losses and an R_{ON} of 0.4 Ω (max). The MTM15N45, with maximum ratings of 15 A and 450 V, was chosen because the induction heater might be operated from either 115 or 220 V sources. A modest heatsink is required because 100 W is dissipated in the power FETs at a full output power of 500 W.

ULTRASONIC TRANSCEIVER

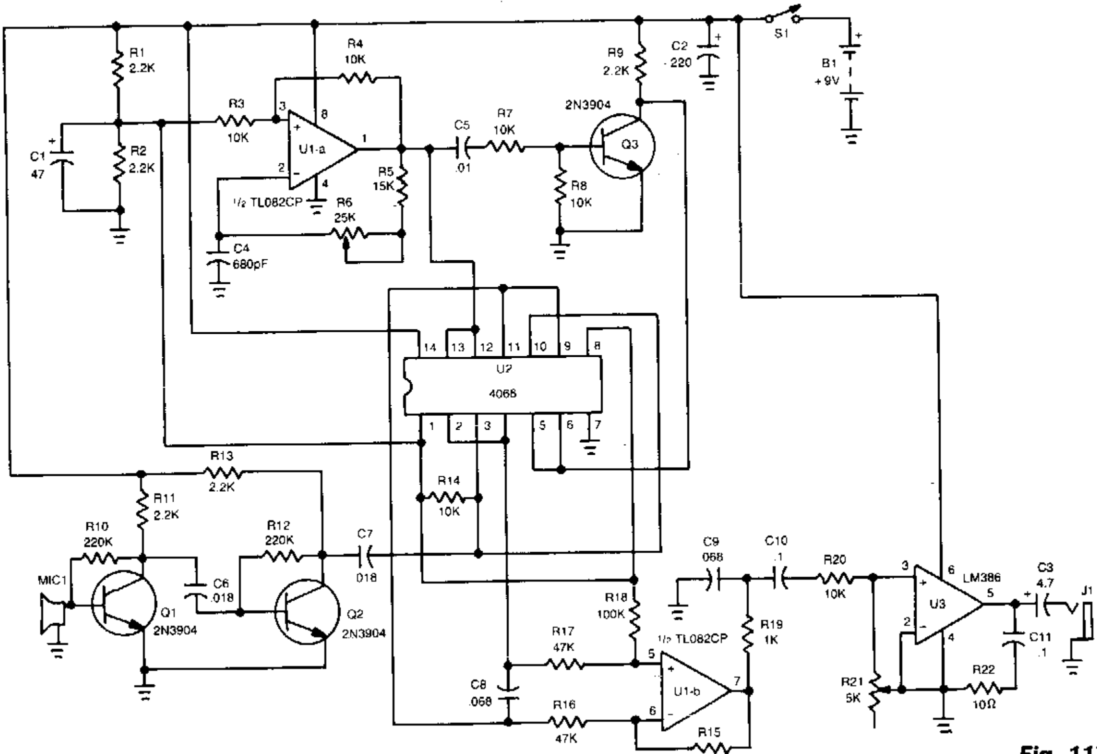


ELECTRONIC DESIGN *Fig. 117-8*



This ultrasonic transmit/receive circuit operates at 40 kHz. Control resistor R5 adjusts the frequency for best performance with the transducers used.

ULTRASONIC RECEIVER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 117-9*

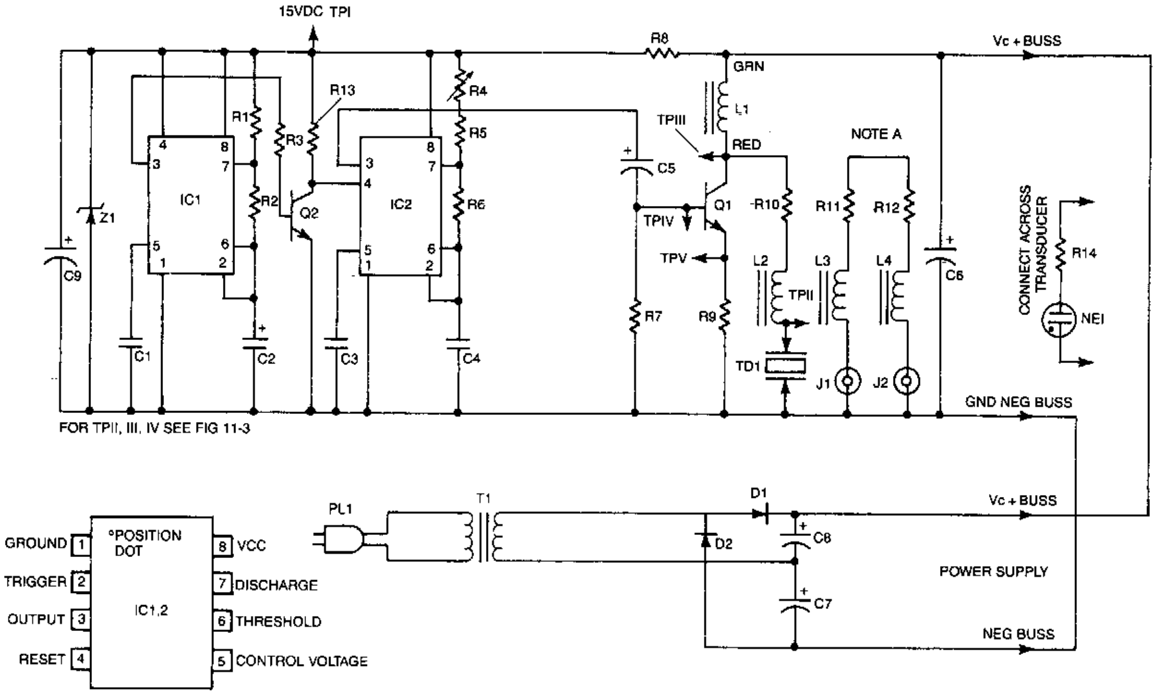
The piezo speaker, MIC1, picks up the incoming ultrasonic signal and feeds it to the base of Q1. The two-transistor booster amplifier, Q1 and Q2, raises the signal to a level that is sufficient to drive one input of this most unusual mixer circuit.

Integrated circuit U2, a quad bilateral switch, functions as an extremely clean balanced-mixer circuit for the superheterodyne receiver. Integrated circuit U1a, 1/2 of a dual op amp, is connected in a variable-frequency square-wave oscillator circuit. Resistors R5, R6, and capacitor C4 determine the frequency and tuning range of the oscillator.

The oscillator's square-wave output is fed along two paths. In one path, the output of U1a is input to pins 12 and 13 of U2. In the other path, the signal is fed to the base of Q3, which is configured as an inverter. The inverter outputs a signal that is 180° out-of-phase with the input signal. The inverted output of Q3 is then fed to U2 at pins 5 and 6. There, the two input signals, the ultrasonic input from MIC1 and the oscillator output, are mixed. The mixing of the ultrasonic input and the square-wave signal produces an audible product that is fed to the input of a differential amplifier, U1b, the second half of the dual op amp, which has a voltage gain of two. The output of U1b at pin 7 is filtered by R19 and C9 to remove the high-frequency content of the mixed signal.

Only the difference frequency is important; the sum frequency, the incoming ultrasonic signal added to the oscillator frequency, is too high for the human ear to hear. The sum frequency is removed by R19 and C9 to produce a clean output signal to feed power-amplifier U3. Resistor R21 functions as the circuit's volume control.

ULTRASONIC-PULSED PEST CONTROLLER



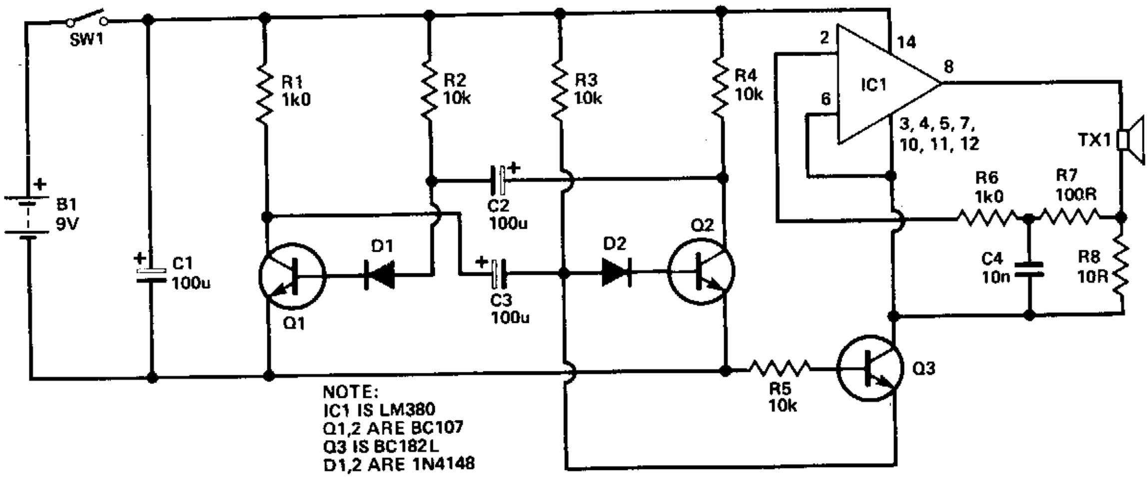
Note a-R11, L3, J1 and R12, L4, J2 are for extra station transducers. Each station should be a similar cabinet with transducer mounted as shown in Fig. 11-3. Use of interconnecting cable strain-reliefs, jack or whatever, left up to builder. Station transducers connected same as TD1.

Use sleeving or tubing on any interconnecting wires that are potentially close to one another or on wire jump bridges. Use component leads wherever possible.

TAB BOOKS ***Fig. 117-10***

IC2 forms a stable oscillator whose frequency and pulse width is determined by the values of R4, R5, R6, and C4. R4 is made adjustable for precise frequency setting. The output of IC2 is pin 3, which is capacitively coupled to the base of Q1. L1 acts as a high-impedance choke to the signal, while allowing the collector of Q1 to be dc-biased. Q1 amplifies the positive pulses from IC2 and step drives the series resonant combination of L2 and TD1. Resistor R10 serves to broaden the response of this resonant circuit. L2 and the inherent capacity of the transducer, TD1, forms a resonant circuit at around 23 kHz. It is usually found that most rodents are bothered when the signal is pulsed with the off exceeding the on time. This timing is accomplished via timer IC1 and timer inverter Q2. IC1 is free running and its periods are determined by R1, R2, and C2 to be approximately two seconds off and two seconds on. The periods are inverted via Q2 and used to gate pin 4 of IC2, the frequency oscillator, turning it on for two seconds and off for three seconds. The power supply is a conventional voltage doubler with a zener regulator for the oscillator voltages.

ULTRASONIC PEST CONTROLLER



ELECTRONICS TODAY INTERNATIONAL

Fig. 117-11

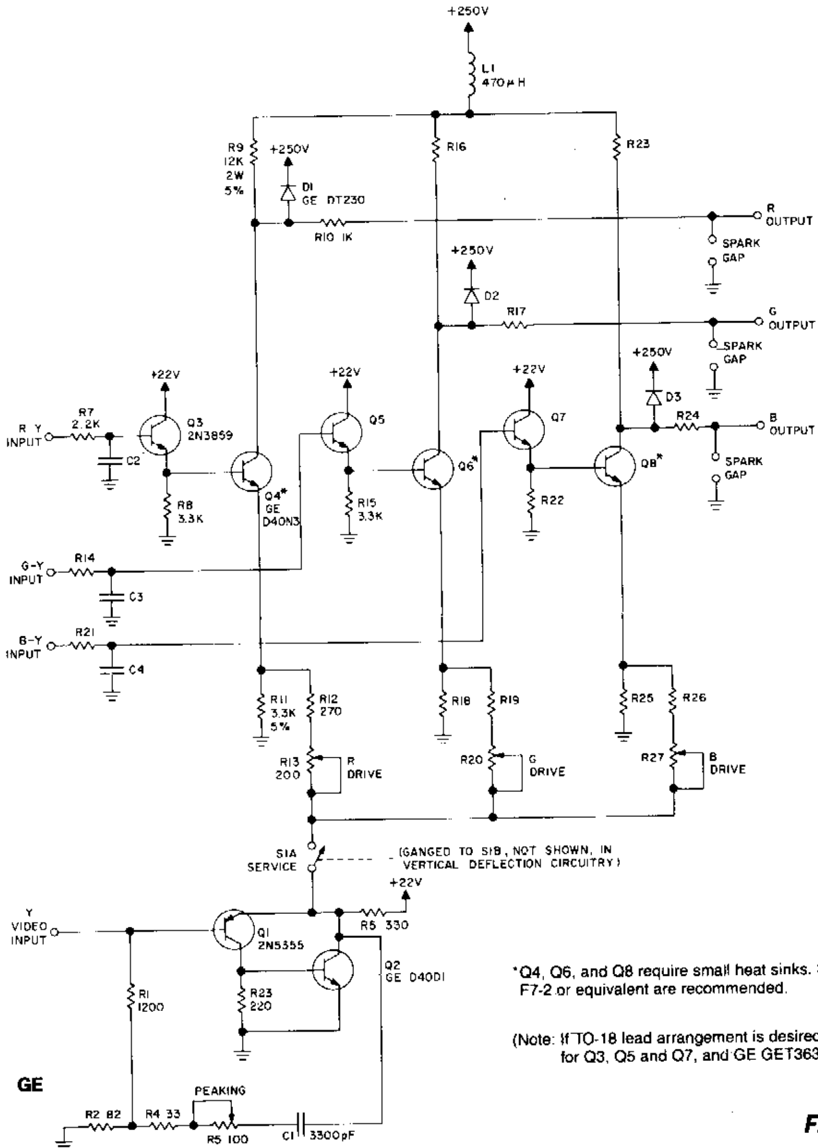
This circuit consists of two basic parts: an oscillator tuned to 40 kHz, and a voltage doubler with pulse generator. The pulses are about 10 ms long and occur 2 – 3 per s to reduce battery drain and increase the annoyance factor for a cat, dog, hedgehog, etc. The voltage doubling action increases the available output power for any given battery voltage.

118

Video Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

RGB VIDEO AMPLIFIER



*Q4, Q6, and Q8 require small heat sinks. Staver type F7-2 or equivalent are recommended.

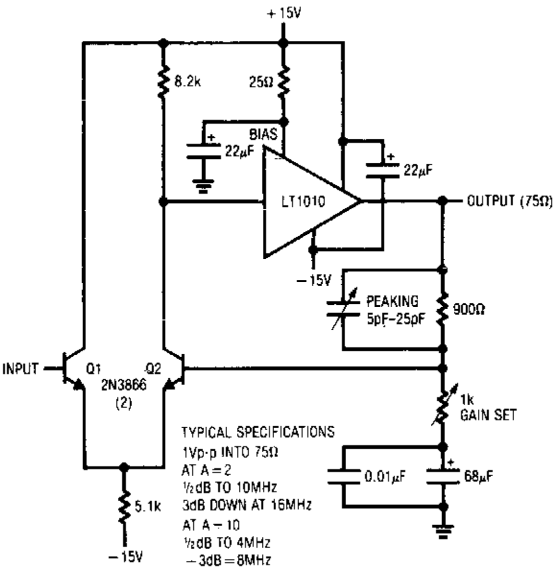
(Note: If TO-18 lead arrangement is desired, use 2N5825 for Q3, Q5 and Q7, and GE GET3638A for Q1.)

GE

Fig. 118-1

Transistors Q1 and Q2 and their associated components provide: a low-impedance output with the necessary power to drive the output stages, give increased gain to high frequencies, and peaking the video for enhanced transient response. Emitter followers Q3, Q5, and Q7 provides low-impedance drive to output stages, Q4, Q6, and Q8. The output stages, with the color difference signals applied to their bases and the luminance signals to their emitters, perform matrixing. The matrixing results in composite output information, to the picture tube, which contains both luminance and chroma information.

VIDEO LINE DRIVING AMPLIFIER

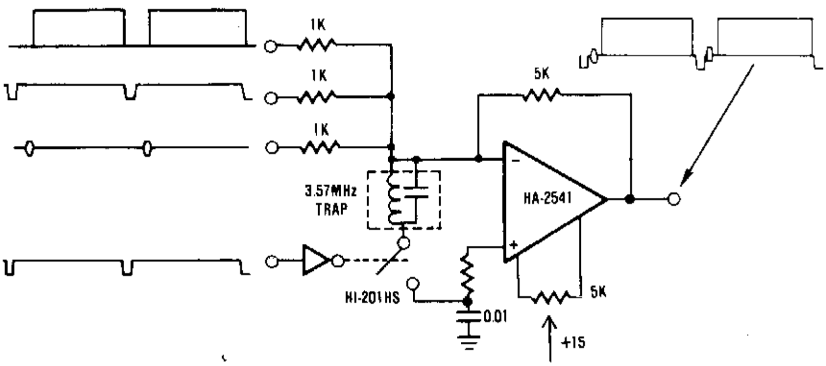


LINEAR TECHNOLOGY CORP.

Fig. 118-2

Q1 and Q2 form a differential stage which single-ends into the LT1010. The capacitively terminated feedback divider gives the circuit a dc gain of 1, while allowing ac gains up to 10. Using a 20-Ω bias resistor, the circuit delivers 1 V pk-pk into a typical 75-Ω video load. For applications sensitive to NTSC requirements, dropping the bias resistor value will aid performance. At $A = 2$, the gain is within 0.5 dB to 10 MHz and the −3 dB point occurs at 16 MHz. At $A = 10$, the gain is flat, within ±0.5 dB to 4 MHz, and the −3 dB point occurs at 8 MHz. The peaking adjustment should be optimized under loaded output conditions.

SUMMING AMPLIFIER/CLAMPING CIRCUIT

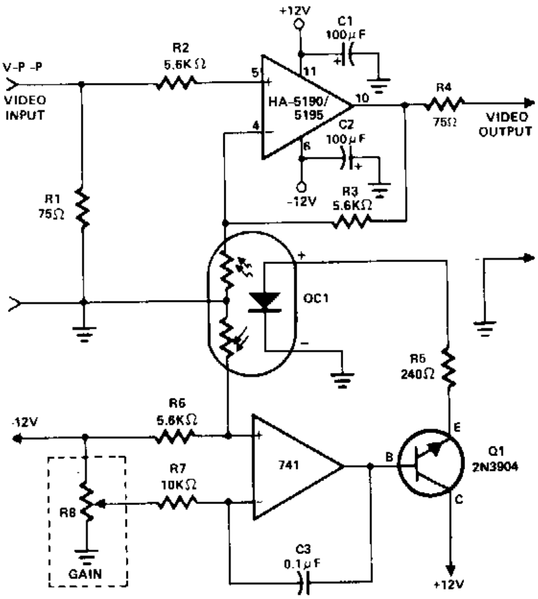


HARRIS

Fig. 118-3

This circuit is a traditional summing amplifier configuration with the addition of the dc clamping circuit. The operation is quite simple; each component—synchronization, color burst, picture information, etc.—of the composite video signal is applied to its own input terminal of the amplifier. These signals combine algebraically and form the composite signal at the output. The clamping circuit, if used, restores the 0-V reference of the composite signal.

DC GAIN-CONTROLLED VIDEO AMPLIFIER



HARRIS

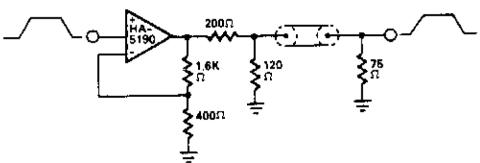
OC1-CLAIREX CLM 8500/2
C1, C2 - TANTALUM TYPES

Fig. 118-4

This amplifier employs a cascaded op amp integrator and transistor buffer, Q1, to drive the gain control element. Except for a simple modification, the HA-5190/5195 stage is connected as a conventional noninverting op amp, and includes input and output impedance matching resistors R1 and R4, respectively, series stabilization resistor R2, and power supply bypass capacitors C1 and C2. The circuit differs from standard designs in that the gain control network includes a photoresistor, part of OC1. The optocoupler/isolator OC1 contains two matched photoresistors, both activated by a common LED. The effective resistances offered by these devices are inversely proportional to the light emitted by the LED. One photoresistor is part, with R3, of the HA-5190/5195 gain network, while the other forms a voltage-divider with R6 to control the bias applied to the integrator noninverting terminal.

In operation, the dc voltage supplied by gain control R8 is applied to the integrator inverting input terminal through input resistor R7. Depending on the relative magnitude of the control voltage, the integrator output will either charge or discharge C3. This change in output, amplified by Q1, controls the current supplied to the OC1 LED through series limiting resistor R5. The action continues until the voltage applied to the integrator noninverting input by the R6—photoresistor gain network is changing, adjusting the op amp stage gain. As the control voltage at R8 is readjusted, the OC1 photoresistances track these changes, automatically readjusting the op amp in accordance with the new control voltage setting.

75-Ω VIDEO PULSE AMPLIFIER



HARRIS

Fig. 118-5

HA-5190 can drive the 75-Ω coaxial cable with signals up to 2.5 V pk-pk without the need for current boosting. In this circuit, the overall gain is approximately unity because of the impedance matching network.

VIDEO GAIN BLOCK

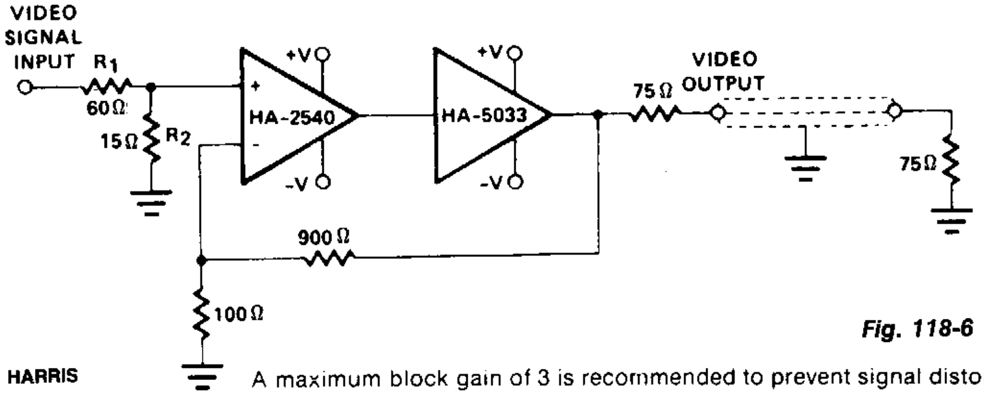


Fig. 118-6

HARRIS

A maximum block gain of 3 is recommended to prevent signal distortion.

This configuration utilizes the wide bandwidth and speed of HA-2540, plus the output capability of HA-5033. Stabilization circuitry is avoided by operating HA-2540 at a closed loop gain of 10, while maintaining an overall block gain of unity. However, gain of the block can be varied using the equation:

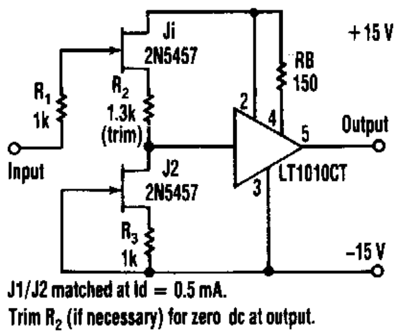
$$\frac{V_{OUT}}{V_{IN}} = 5\,\frac{R2}{(R1 + R2)}$$

$$\text{where } R1 + R2 = 75\ \Omega$$

A maximum block gain of 3 is recommended to prevent signal distortion.

This circuit was tested for differential phase and differential gain using a Tektronix 520A vector scope and a Tektronix 146 video signal generator. Both differential phase and differential gain were too small to be measured.

LOW-DISTORTION VIDEO BUFFER



ELECTRONIC DESIGN

Fig. 118-7

This buffer amplifier's overall harmonic distortion is a low 0.01% or less at 3-V rms output into a 500-Ω load with no overall feedback. The LT1010CT offers a 100 V/μs slew rate, a 20 MHz video bandwidth, and 100 mA of output. A pair of JFETs, J1 and J2 are preselected for a nominal match at the bias level of the linearized source-follower input stage, at about 0.5 mA. The source-bias resistor, R2, of J1 is somewhat larger than R3 so that it can drop a larger voltage and cancel the LT1010CT's offset. J1 and J2 provide an untrimmed dc offset of ±50 mV or less. Swapping J1 and J2 or trimming the R2 value can give a finer match.

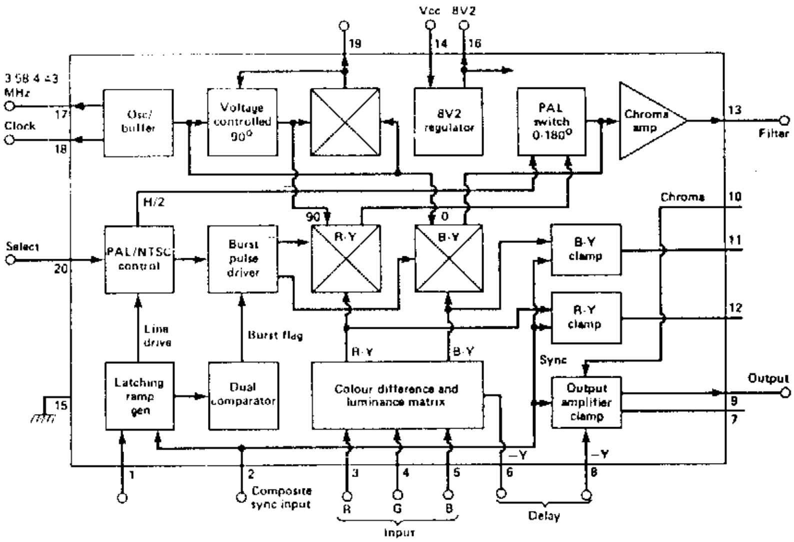
The circuit's overall harmonic distortion is low: 0.01% or less at 3-V rms output into a 500-Ω load with no overall feedback. The circuit's response to a ±5 V, 10 kHz square-wave input, band-limited to 1 μs, has no overshoot. If needed, setting bias resistor R_B lower can accommodate even steeper input-signal slopes and drive lower impedance loads with high linearity. The main trade-off for both objectives is more power dissipation. A secondary trade-off is the need for retrimming the source-bias resistor, R2.

119

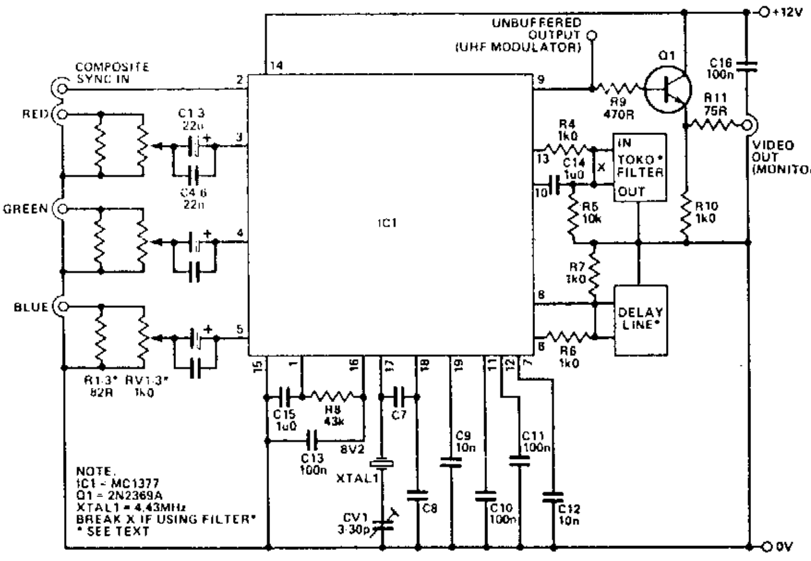
Video Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

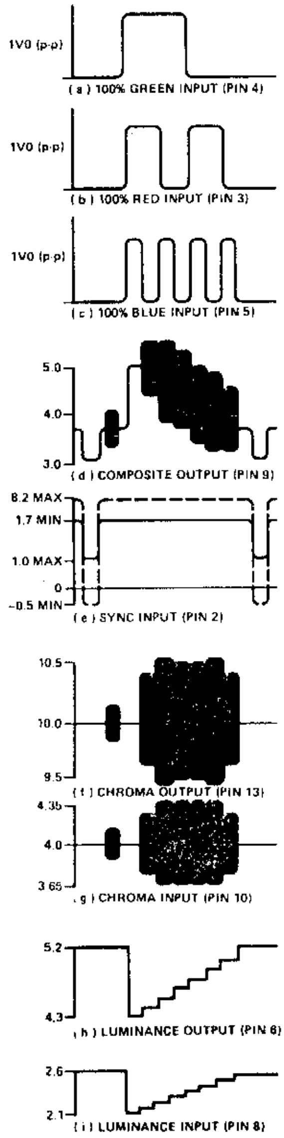
RGB-COMPOSITE CONVERTER



BLOCK DIAGRAM OF THE MC1377P ENCODER IC.



CIRCUIT DIAGRAM OF THE CONVERTER.



The signals that should appear at the test points around the chip.

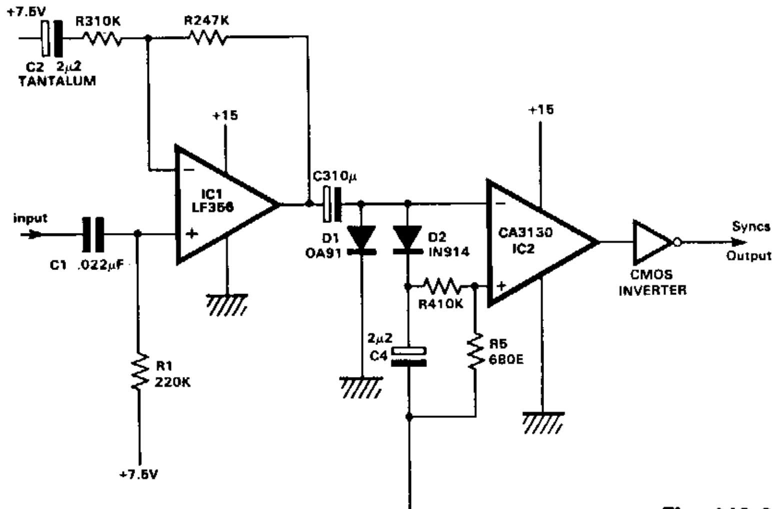
ELECTRONICS DIGEST *Fig. 119-1*

The incoming RGB inputs are terminated with resistors R1, R2, and R3 and potentiometers RV1, RV2, and RV3. These provide input impedances of approximately 75 Ω. The presets should be adjusted to provide a maximum input of 1 V pk-pk into the MC1377. The inputs are ac-coupled into the encoder; the large value capacitor is required for the 60 Hz field component.

The Colpitts oscillator for the color burst is formed around pins 17 and 18. About 0.5 V pk-pk should appear on pin 17 and 0.25 V rms into pin 18 with the oscillator components removed. The incoming composite sync signal at pin 2 should be negative-going. The device will accept CMOS and TTL directly. If it is necessary to ac-couple the sync, then a pull-up to 8.2 V is required—a regulated 8.2 V is provided on pin 16.

From the composite sync input, the MC1377 generates a ramp which it uses to provide the burst gate pulse. The slope of this ramp can be varied by a potentiometer on pin 1. However, a preset value, shown as 43 KΩ, is usually sufficient. The chrominance filter should be fitted between pins 13 and 10. If the filter is not used, a compensatory potential divider should be fitted (both are shown). We used a prealigned Toko bandpass filter centered on 4.43 MHz. If the chroma filter is fitted, the delay through it, 400 ns, has to be compensated for by a luminance delay line between pins 6 and 8. This line is shorted out if the filter is not fitted. The composite video output from the IC is buffered to provide a low-impedance drive for a monitor, or it can be applied directly to a UHF modulator commonly used in computers.

SINGLE-SUPPLY WIDE-RANGE SYNC SEPARATOR



ELECTRONIC ENGINEERING

Fig. 119-2

This circuit extracts the sync pulses from a video signal over a wide range of amplitudes and operates a single +15 V supply. IC1 buffers and amplifies the incoming signal and applies it via C3 to the peak detector, consisting of D2 and C4. It is also applied to one input of a comparator, IC2. The other input of IC2 is set at a voltage corresponding to about 0.065 of the peak video amplitude, by the divider R4/R5. The trigger points of IC2 are set near the bottom of the sync pulses which help prevent spurious noise. These resistors also leak across C4, so they must be chosen as a compromise between excessive ripple and speed of response to falling signal levels. The IC2 output swings between 0 and 15 V and is conveniently CMOS compatible, but further buffering is advisable, hence the CMOS inverter. Maximum input amplitude is set by saturating IC1's output. The minimum acceptable level is set by the forward voltage drop of the dc restoring clamp D1, which should be either a germanium (as shown) or a Schottky diode.

CHROMA DEMODULATOR WITH RGB MATRIX

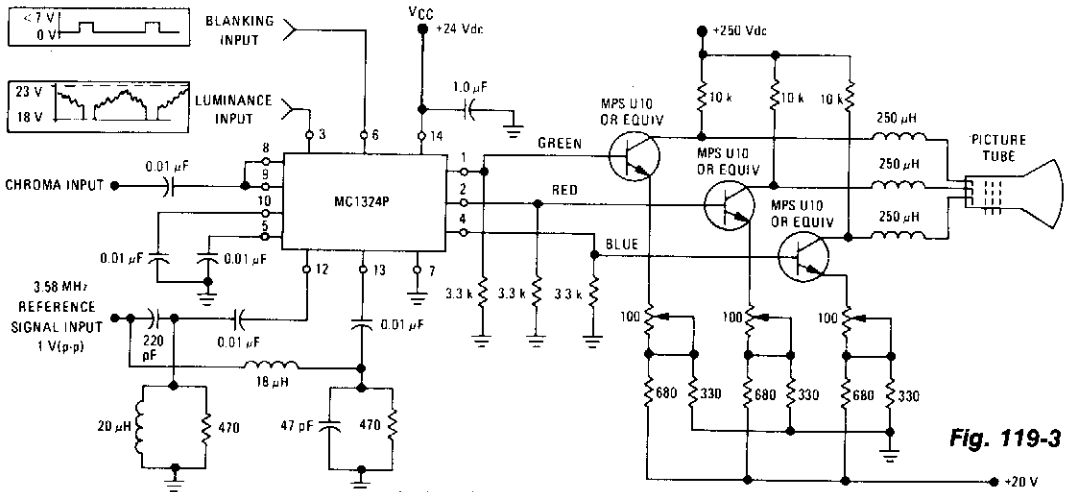
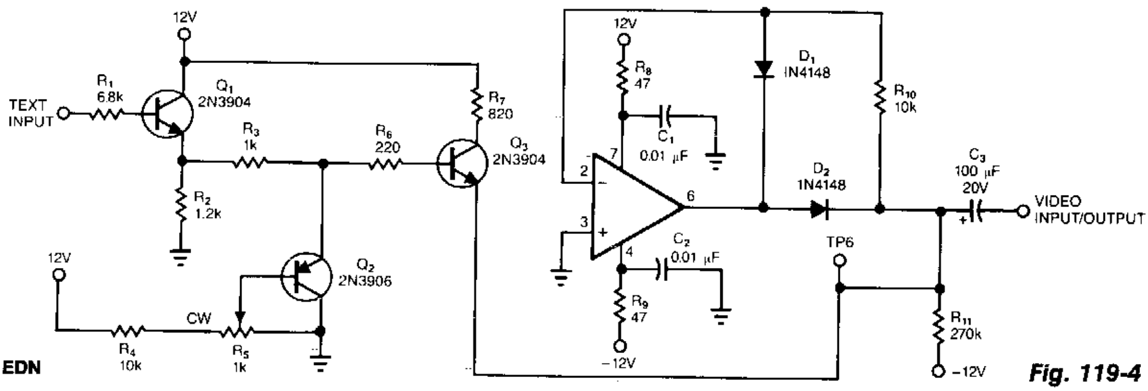


Fig. 119-3

A typical application is given above to indicate the requirements and output functions of this chroma demodulator.

The MC1324 provides chroma demodulation recovering recooling the R, G, and B signals to drive video amps for each color difference signal. The luminance signal and chrominance signal are matrixed to get the R, G, and B signals.

COMPOSITE-VIDEO SIGNAL TEXT ADDER

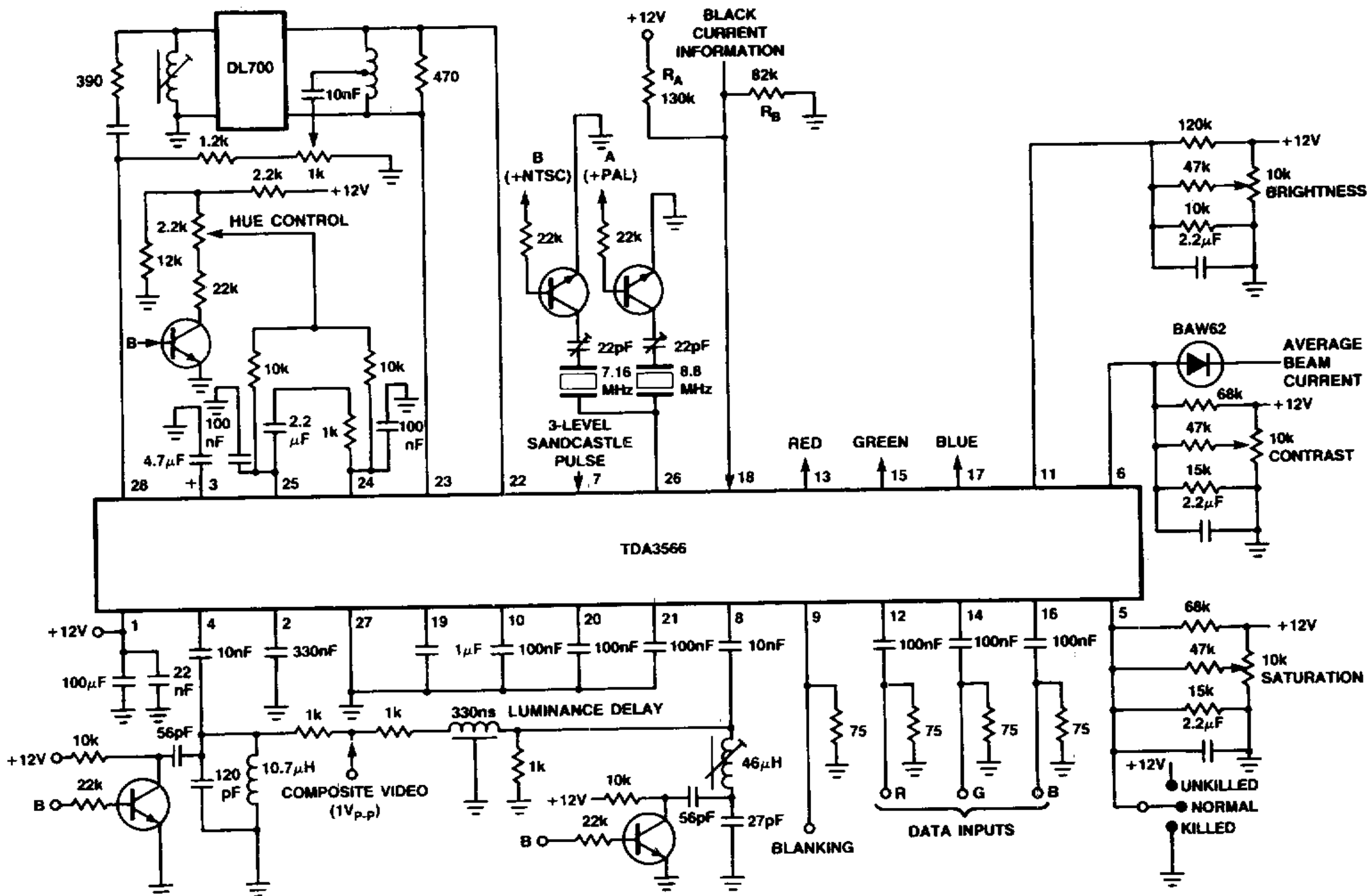


EDN

Fig. 119-4

This circuit shows a simple way to add text information to a composite-video signal that might be floating at some indeterminate dc level. The text generator and composite-video source must have the same sync signal. The video-input and -output signals share the same terminal. C3 couples the video signal to the output of a rectifier circuit that is based on a subvideo-speed op amp. A faster op amp would clamp on individual sync pulses rather than on the video waveform's average value, as is desired. R11 serves as a pulldown resistor and feedback resistor R10 ensures that TP6 remains at ground level. Emitter follower Q1 buffers the text signal, and R5 serves as a gain control. A simple clamp circuit, Q2, is sufficient for regulating amplitude, because the text signal contains no gray-scale levels. Q3 couples the text signal into the op-amp clamp circuit.

PAL/NTSC DECODER WITH RGB INPUTS

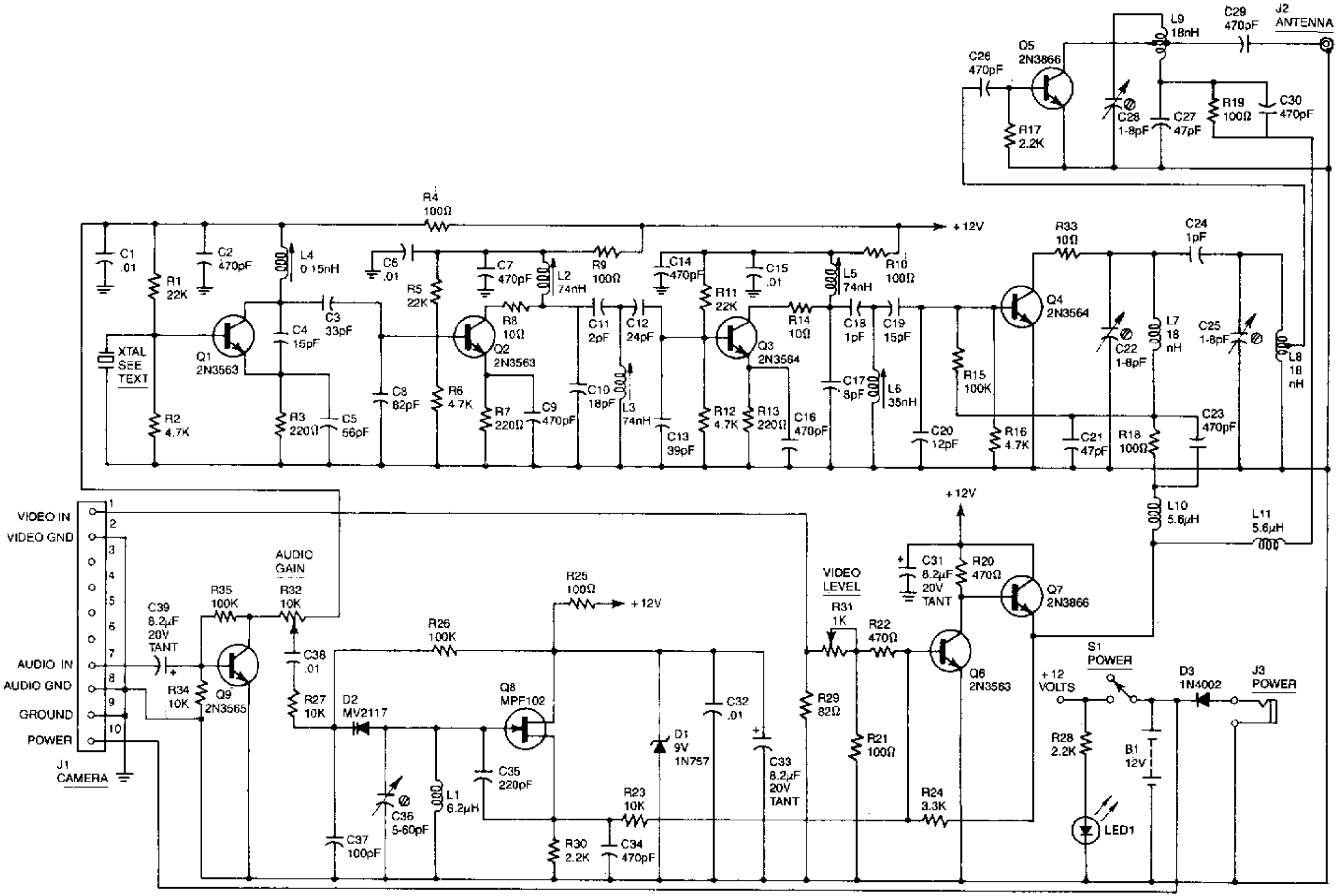


SIGNETICS

Fig. 119-5

This circuit shows the TDA3566 for a PAL/NTSC Decoder.

WIRELESS VIDEO CAMERA LINK

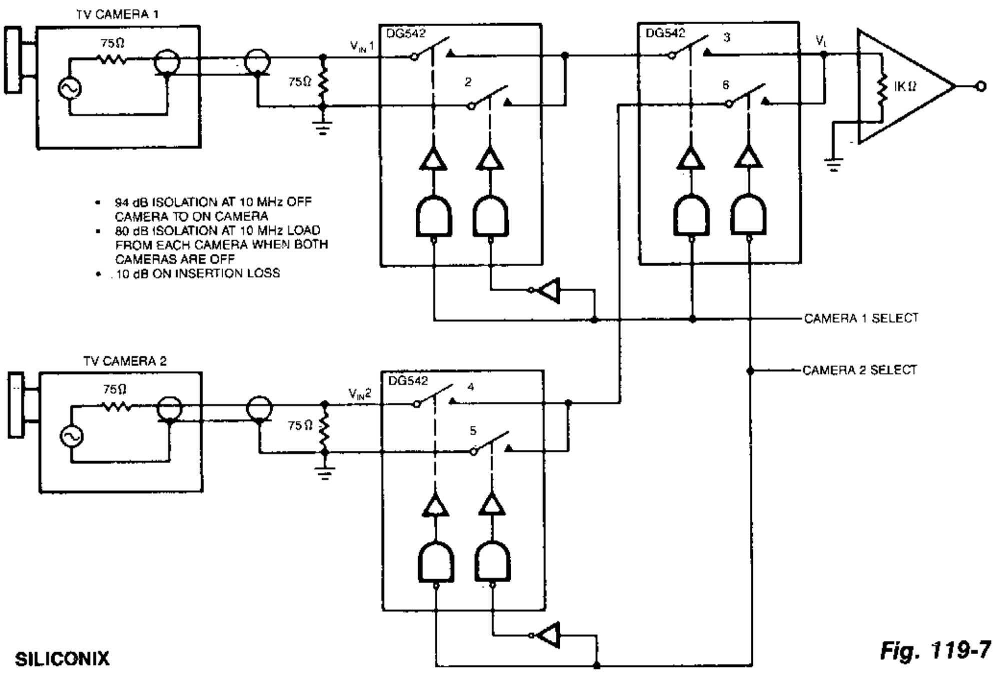


Reprinted with permission from Radio-Electronics Magazine, February 1986. Copyright Gernsback Publications, Inc., 1986.

Fig. 119-6

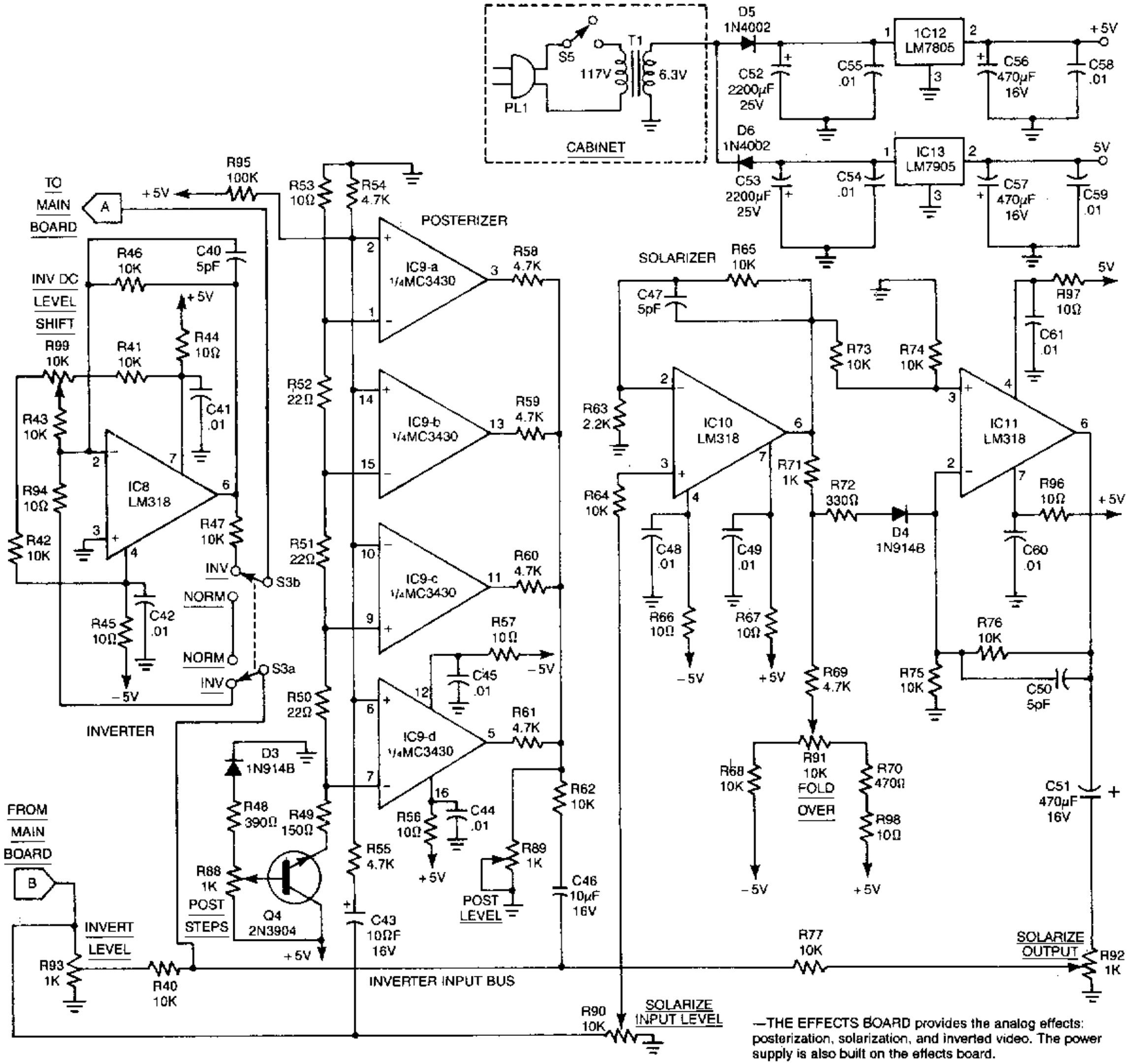
This high-performance video-camera link transmits signals from your video camera to your VCR, or from your VCR to TVs throughout your home. The first stage of the rf chain is a crystal-controlled oscillator, Q1, with a frequency of 60 to 65 MHz, which is one-eighth of the final output frequency. The oscillator produces a signal of about +6 dBm (4 mW) that drives three stages of frequency doublers. The combined action of those doublers multiplies the input frequency by eight for a final output frequency of (nominally) 500 MHz. Double-tuned circuits are used between each stage to help reduce spurious outputs that might cause unwanted interference. The video input signal from your VCR, video camera, etc. drives a video modulator, Q6 and Q7, that adds the video signal to the +12 V line supplying power to the final doubler, Q4, and the output amplifier, Q5. That method of modulation is similar to the way a conventional AM-radio transmitter is modulated. The video modulator has a nominal bandwidth of five MHz. The audio input is applied to Q8, which operates as a VCO running at a nominal frequency of 4.5 MHz to produce the modulated sound carrier. For simplicity, Q8 is a free-running oscillator, since the ±25 kHz frequency deviation that is required would be very difficult to produce at that frequency with a crystal-controlled oscillator. Besides, most TV sound systems will accept a ±10 kHz error in the sound-carrier frequency without producing undue distortion, and that greatly simplifies the circuitry required. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

VIDEO SWITCH WITH VERY HIGH OFF ISOLATION



SILICONIX

Fig. 119-7



—THE EFFECTS BOARD provides the analog effects: posterization, solarization, and inverted video. The power supply is also built on the effects board.

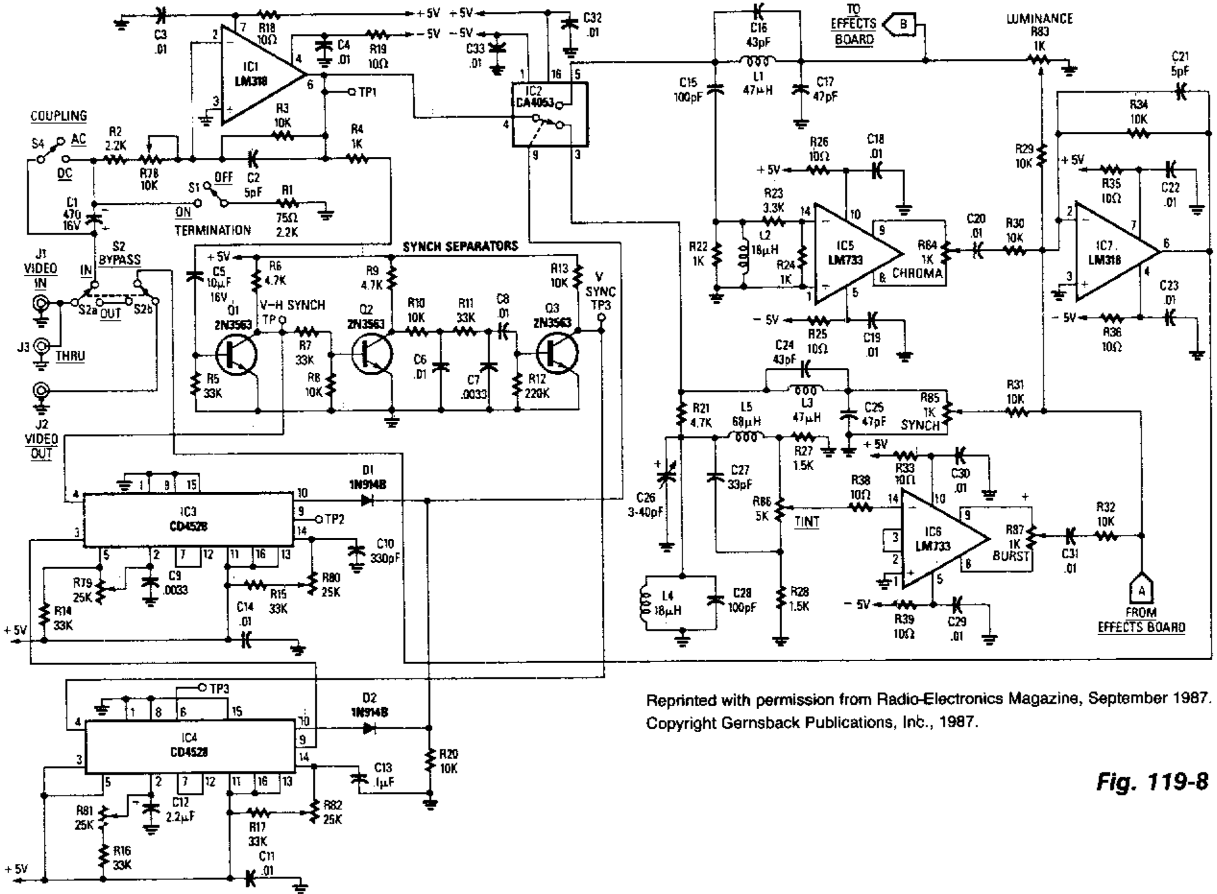
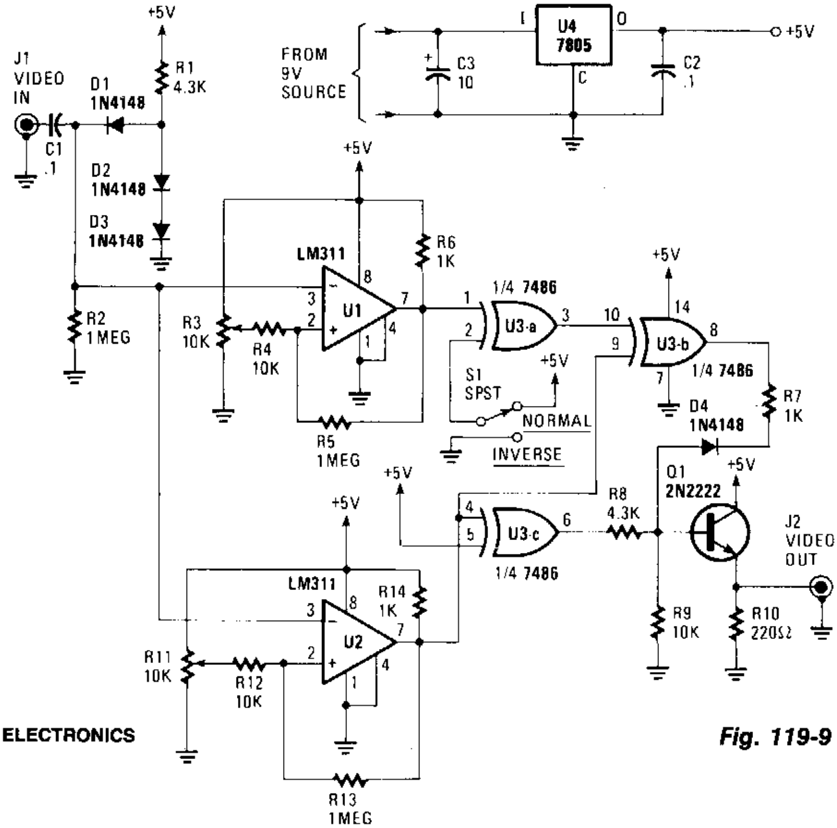


Fig. 119-8

This system consists of two parts. The main board dissects the video signal and provides independent level control for burst, chroma, luminance, and sync signals, as well as phase and polarity. The video signal is reassembled in the output in corrected or modified format as required by the user. The effects board produces luminance inversion or can generate discrete luminance steps (posterization) or a nonlinear gray scale (solarization) to achieve simulation of photographic effects commonly seen in various special-effect photographic processes. The kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

PICTURE FIXER/INVERTER



HANDS-ON ELECTRONICS

Fig. 119-9

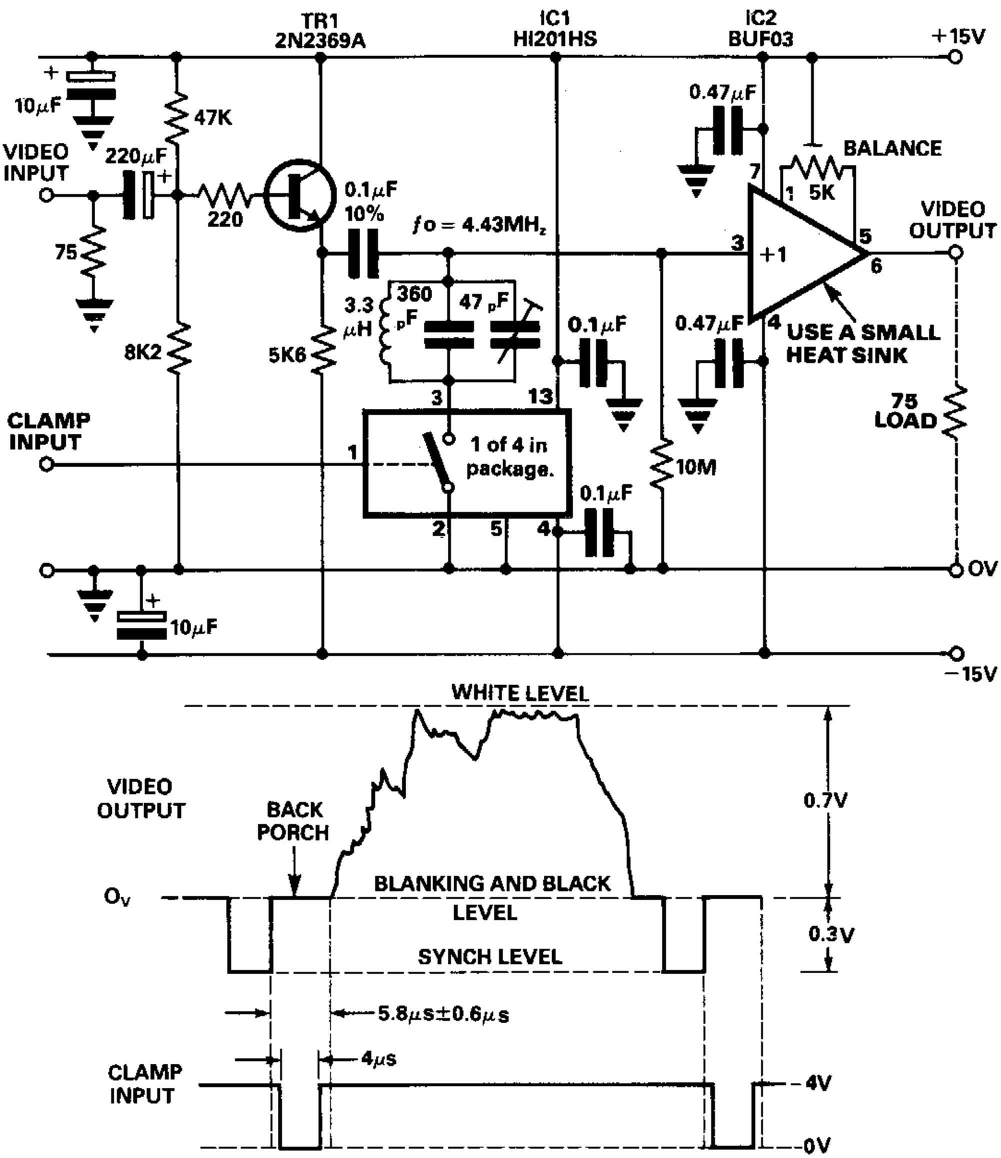
The circuit will accept a video signal, separate the sync pulses, invert the video, and add new video to the old sync pulses.

The video signal is brought in through J1 and applied to a clamping circuit consisting of C1, D1, D2, D3, R1, and R2. The clamp circuit forces all of the sync pulses to align with the same dc voltage level. With the video voltage clamped, the trip points of the comparators that follow can be set with trimmer resistors R3 and R11. The resistors will not have to be readjusted. One comparator, U1, is adjusted to change states with a change in either video or sync-pulse levels. The other comparator, U2, is adjusted to trip on changes of sync-pulse levels only.

The output of U1 now consists of a logic level, 0 to +5 V, signal that contains both sync pulses and video. The composite signal is coupled to an EXCLUSIVE-OR gate, U3a, where it is either inverted or not inverted, depending upon the position of switch NORM/REV S1. The output, at pin 3 of U3a, is next sent to U3b. There the composite signal is combined with the sync-pulse only signal from U2. The EXCLUSIVE-OR action of U3b cancels out the sync pulses, leaving only video at the IC's output.

Since the sync pulses are inverted as they pass through U2, they must be inverted once more before being combined with the video signal. That final inversion is performed by U3c, and that device's output is combined with that of U3b via D4, R7, R8, and R9. The newly combined signal is buffered by emitter-follower Q1, and sent to the outside world via J2. The circuit can be powered by a 9- to 12-V wall-mount power supply. The supply voltage is regulated down to 5 V by U4.

VIDEO DC RESTORER

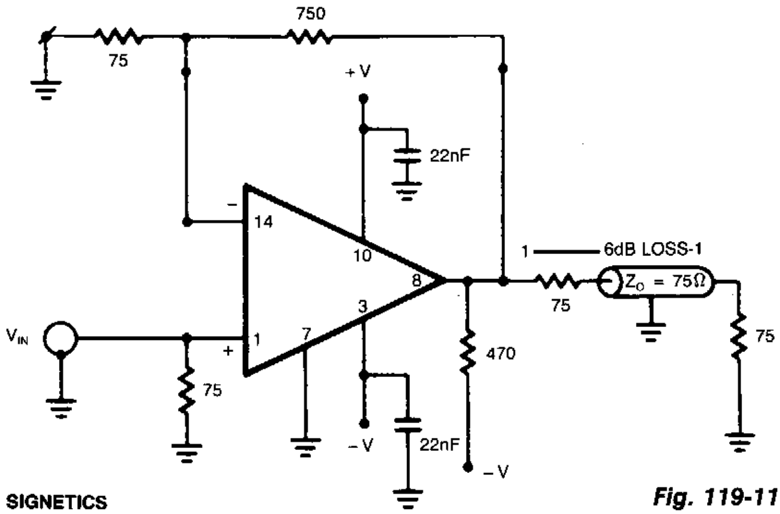


ELECTRONIC ENGINEERING

Fig. 119-10

The main requirement for efficient dc restoration is to provide a short time-constant during the clamp period, with a long time-constant during the active line time. The switch within the Harris HI201HS has an on resistance of 30 Ω and the PMI buffer, BUF03, has an input resistance of 50 MΩ. The tuned circuit presents a high impedance at the 4.43-MHz color subcarrier frequency so that the color-burst signal is retained if the video signal contains this information.

COLOR VIDEO AMPLIFIER

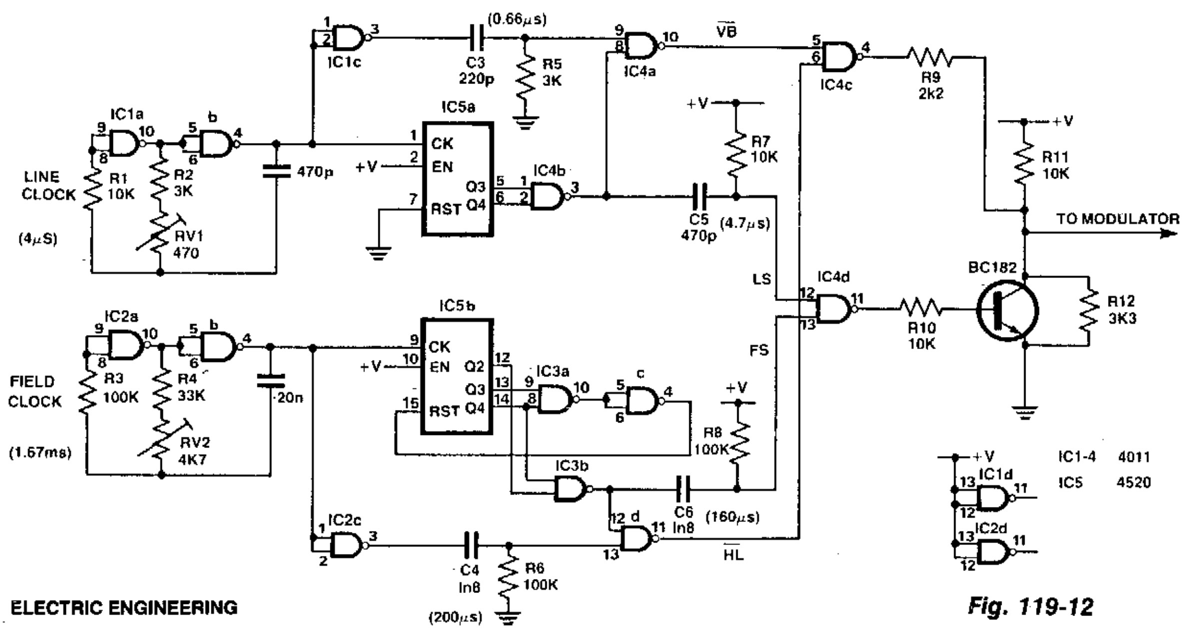


SIGNETICS

Fig. 119-11

The NE5539 wideband op amp is easily adapted for use as a color video amplifier. The gain varies less than 0.5% from the bottom to the top of the staircase. The maximum differential phase is approximately +0.1°. The amplifier circuit was optimized for a 75-Ω input and output termination impedance for a gain of approximately 10 (20 dB).

COLOR TV CROSSHATCH GENERATOR

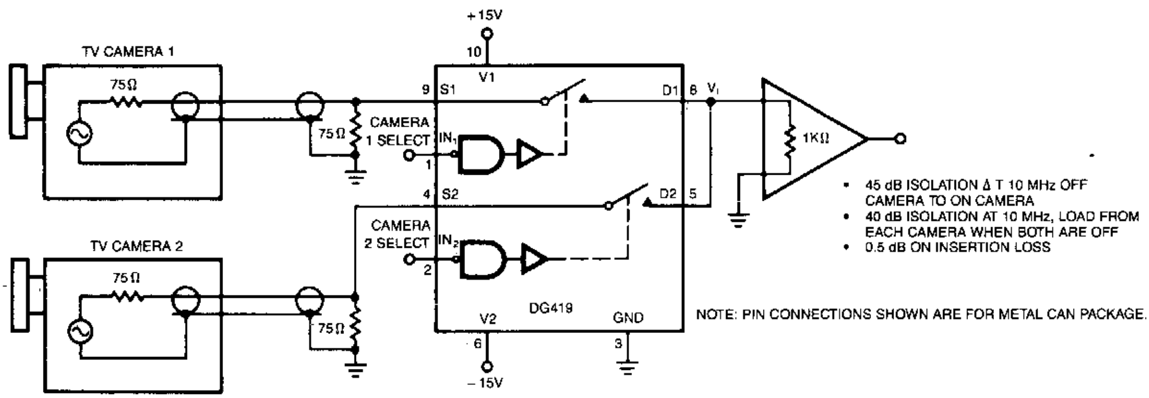


ELECTRIC ENGINEERING

Fig. 119-12

This circuit provides a simple, low-cost crosshatch generator for convergence and geometry adjustments on color TVs. The generator is driven by two clocks, one for the horizontal drive, IC1ab, and one for the vertical drive, IC2ab. The clock outputs are applied to the two binary counters contained in IC5 which generate the line and field sync pulses and respective blanking periods. Line clock pulses, buffered by IC1c, are differentiated by C3/R5 to produce the vertical bars. These bars are gated by IC4a which suppresses the bars during the line blanking period produced by the coincidence of Q3, Q4 outputs of IC5a, detected by IC4b. This output is also differentiated by C5/R7 to produce the line sync pulse, *LS*. A similar process is used to generate horizontal lines and the field sync, except that in order to give the correct aspect ratio, the count of IC5b is reset at 12, coincidence of Q3 and Q4. In coincidence of Q2, Q4 is used to generate the sync pulse *FS* and the blanking period. The line and field sync pulses, *LS* and *FS*, are combined in IC4d.

GENERAL PURPOSE VIDEO SWITCH

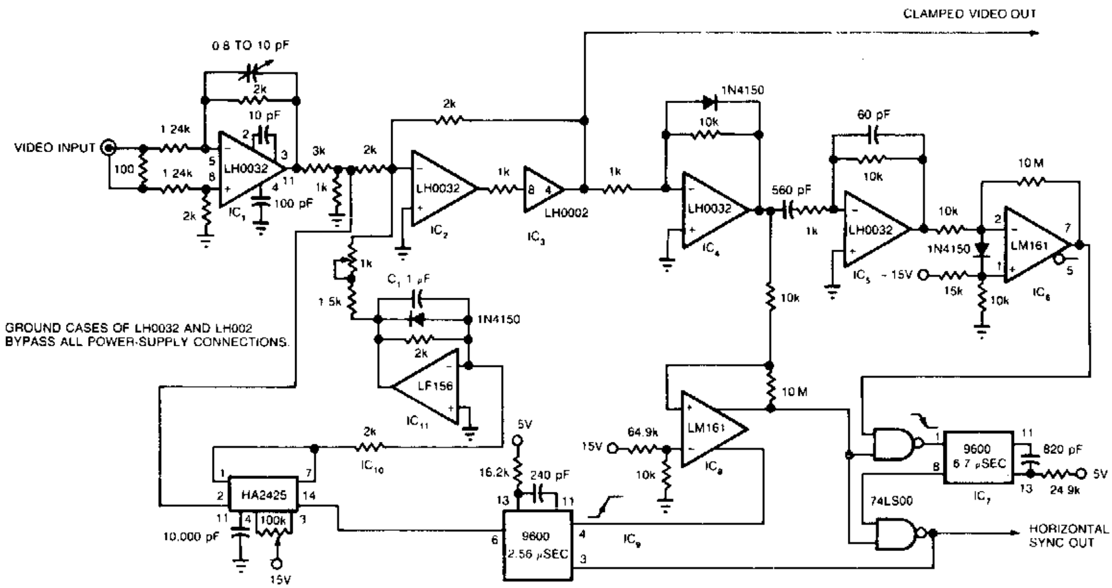


SILICONIX

Fig. 119-13

The circuit shown provides 40-dB isolation at 6 MHz and is good for general purpose video switching.

VIDEO SIGNAL CLAMP

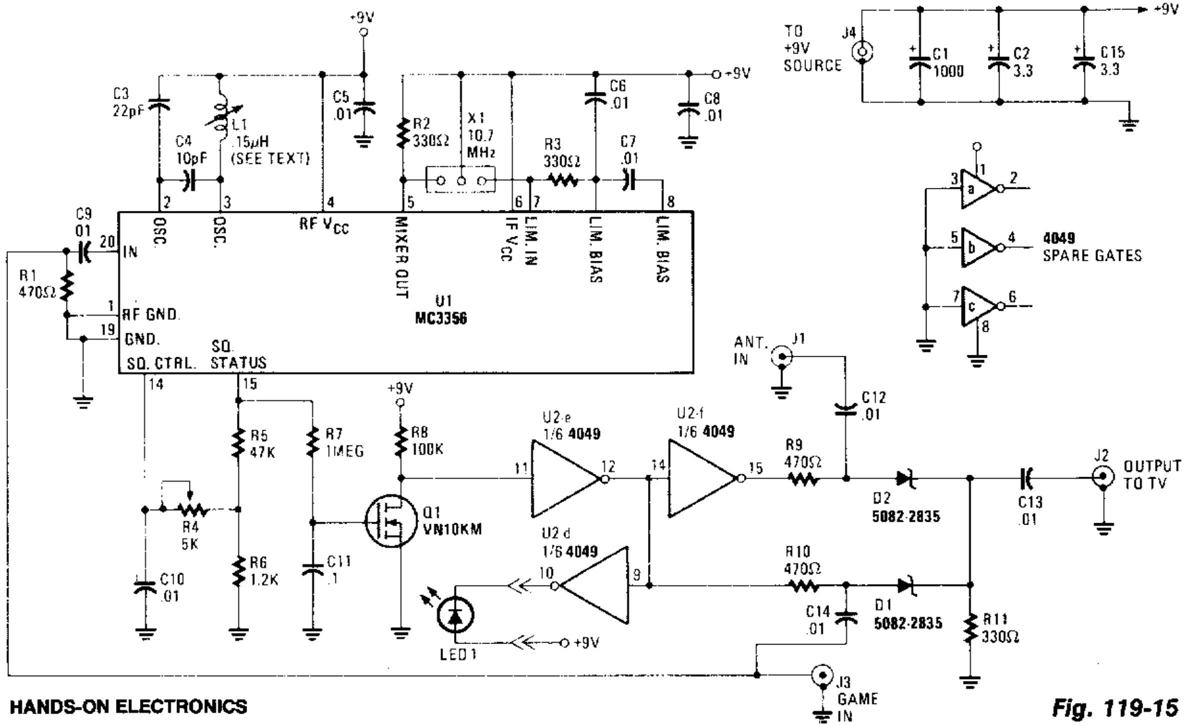


EDN

Fig. 119-14

The circuit uses a track-and-hold amplifier in a closed-loop configuration to clamp the back-porch voltage of a standard video waveform to 0 V. The circuit's outputs include a clamped composite-video signal and a TTL-level horizontal-blanking pulse. Differential input buffer IC1 and the summing amplifier IC2 isolate the input video signal. Clipper IC4 removes the video signal, leaving only the synchronization information. Differentiator IC5 detects the edges of the horizontal blanking pulses and produces pulses that correspond to the leading and trailing edges of the horizontal blanking pulses. IC6 clips these pulses and converts them to a TTL level. IC7 uses these clipped pulses to generate a TTL-level window that, when combined with the horizontal pulse generated by IC8, forms a TTL representation of the original horizontal pulse. This representation is synchronized to the input waveform. IC9 uses the trailing edge of this reconstructed waveform to generate the track pulse for track-and-hold amplifier IC10. IC11 filters IC10's dc output and, after gain adjustment, feeds it back to IC2's summing node.

AUTOMATIC VIDEO SWITCH

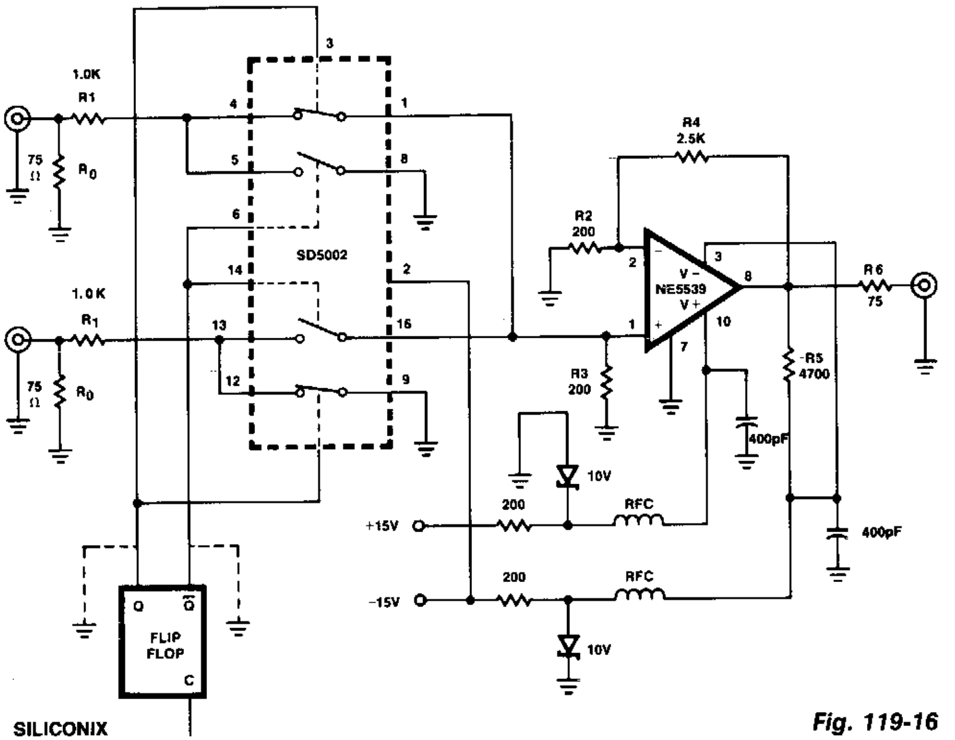


HANDS-ON ELECTRONICS

Fig. 119-15

Turn on a game, computer, videodisc player, or whatever, and its output takes priority over the antenna-derived signal—the antenna is disconnected and the alternate source is fed to the monitor. When the alternate video source is no longer detected, the switch automatically reconnects the antenna. When the rf carrier is detected, a high logic output appears at pin 15 of U1—squelch status. The signal is then buffered to VMOS FET Q1. This FET can drive CMOS inverter/buffer U2, can be an *open-drain* output to drive a relay, or can convert to a 5 V logic level. When pin 15 of U1 becomes high, Q1 turns on, pulling pin 11 of U2e low; in turn, pin 12 becomes high. The output of U2e is fed to U2f, forcing its output, at pin 15, low. When the signal at U2 pin 12 is high, D1 is biased on, allowing the signal at J3 to flow through C14, D1, and out to J2 through C13. When the signal at J3 is removed, U1 pin 15 decreases. That decrease causes U2 pin 12 to decrease and U2 pin 15 to increase. When the signal at U2 pin 15 is high, it biases D2 on, allowing the signals at J1 to flow through C12, D2, and out to J2 through C13.

HIGH-PERFORMANCE VIDEO SWITCH



SILICONIX

Fig. 119-16

This figure shows a one-of-two switch with a summing amplifier. The video source's line can be terminated either externally or internally to switch R0. With this termination resistor, a load change of less than 1 Ω will be *seen* by the source when the switch changes state. For this reason, input isolation amplifiers are not necessary. R4 can be varied to control circuit gain, but should never be less than 1400 Ω since the NE5539 is internally compensated for gain values greater than seven. A value of approximately 2500 Ω for R4 will set circuit gain to near unity. Additionally, the circuit output impedance is set by R6, and R5 sets the output dc offset to near zero.

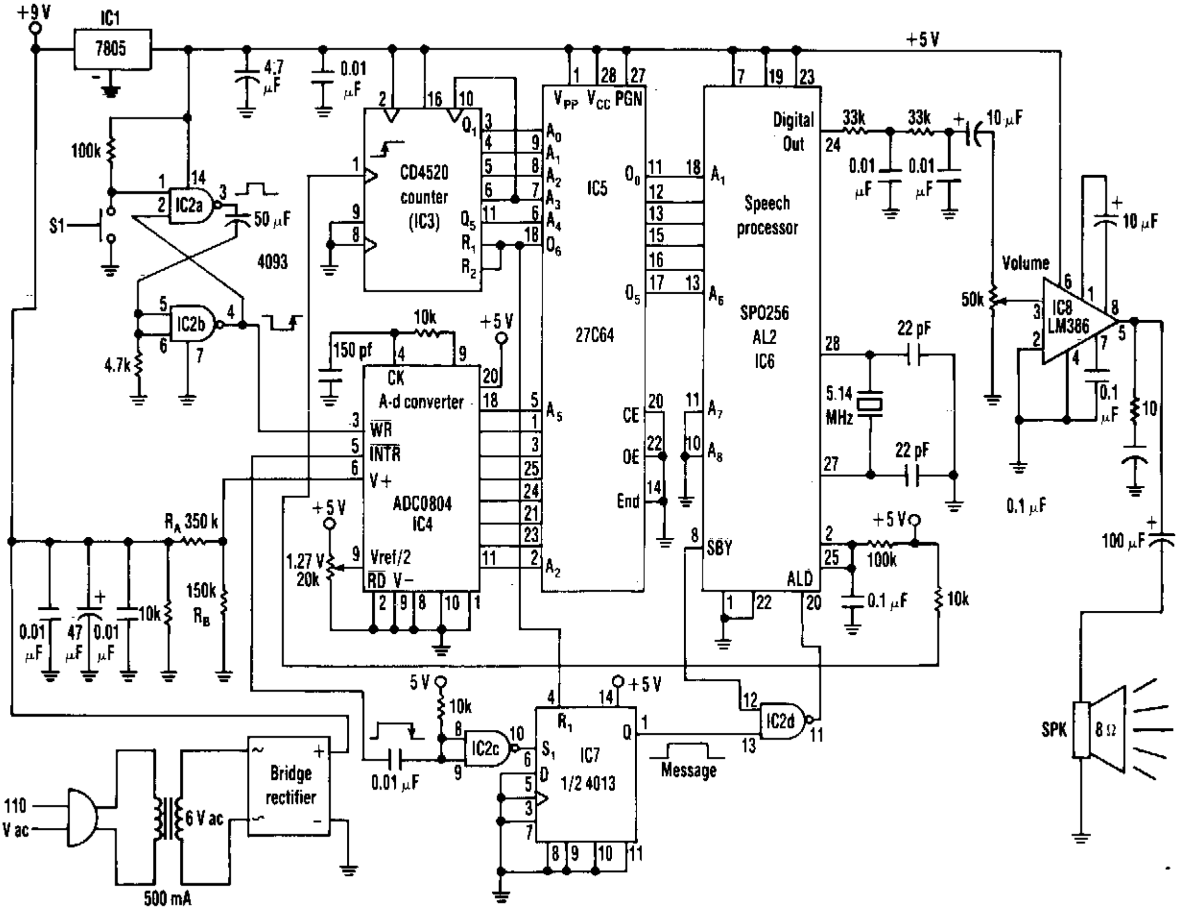
120

Voice Circuits

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Ac Line-Voltage Announcer
Dialed Phone Number Vocalizer
Computer Speech Synthesizer
Allophone Generator
Electronic Voice Substitute

AC LINE-VOLTAGE ANNOUNCER



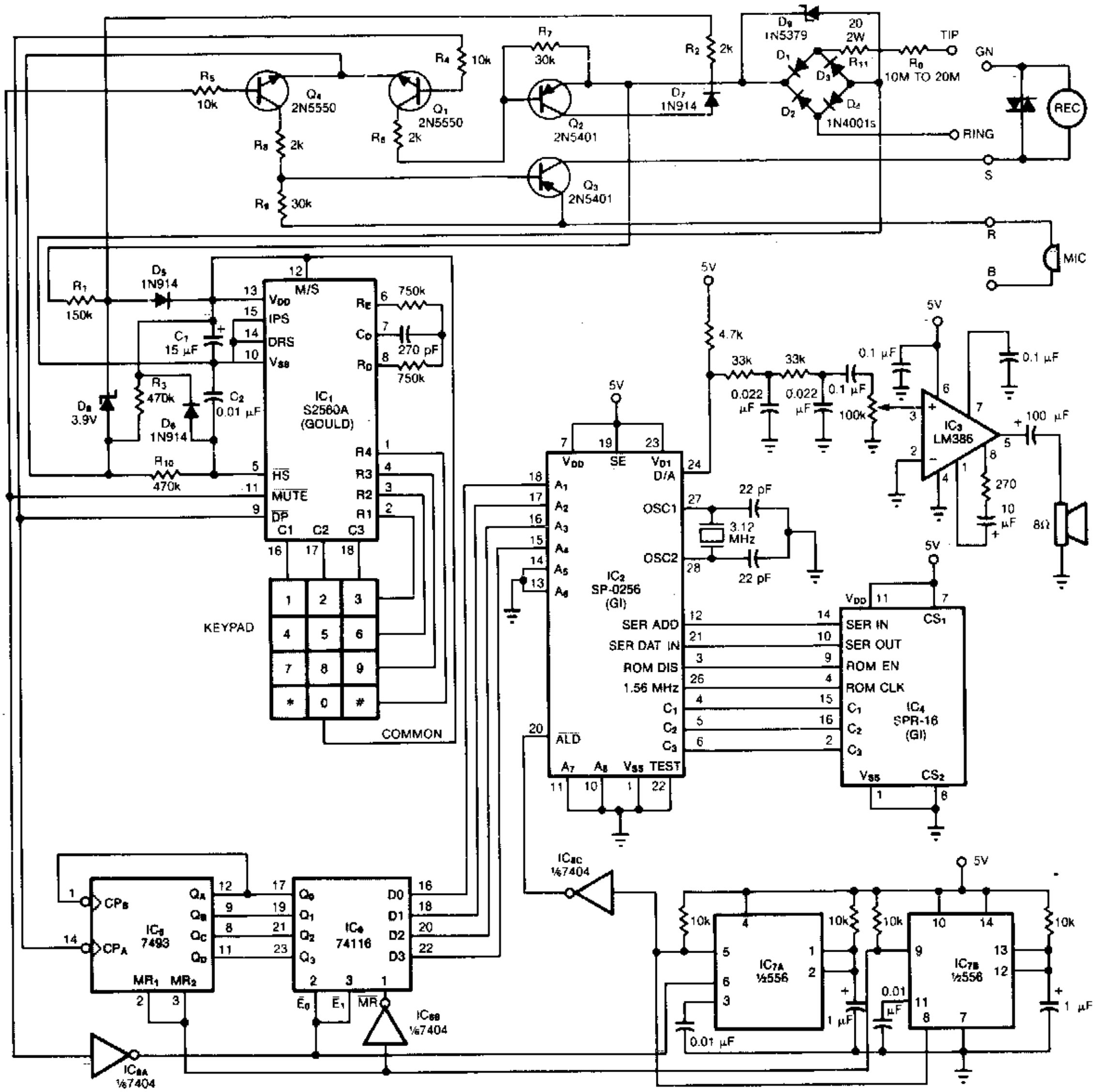
ELECTRONIC DESIGN

Fig. 120-1

The range of this ac-voltage monitor is 100 to 140 Vac, with a resolution of 1 V. The speech processor interprets an 8-bit binary input code from an analog-to-digital converter. The processor's pulse-code-modulated output then passes through a filter and an amplifier before driving the circuit's speaker to vocalize the corresponding number. Each time switch S1 is pressed, the speech-processor program enunciates the monitored voltage readings from 100 to 140 V, depending on the code at the input of a 27C64 EPROM.

The voltage-monitoring circuit consists of a bridge rectifier, filter capacitors, and a 10-KΩ load resistor. A divider, R_A and R_B, limits the input voltage to a maximum 2.55 V. The a/d converter, IC4, then sends the voltage reading to the 27C64 EPROM, IC5. Pressing S1 sends a negative transient pulse to the write, WR, input of the a/d converter, IC4, which initiates a 100-μs conversion process.

DIALED PHONE NUMBER VOCALIZER

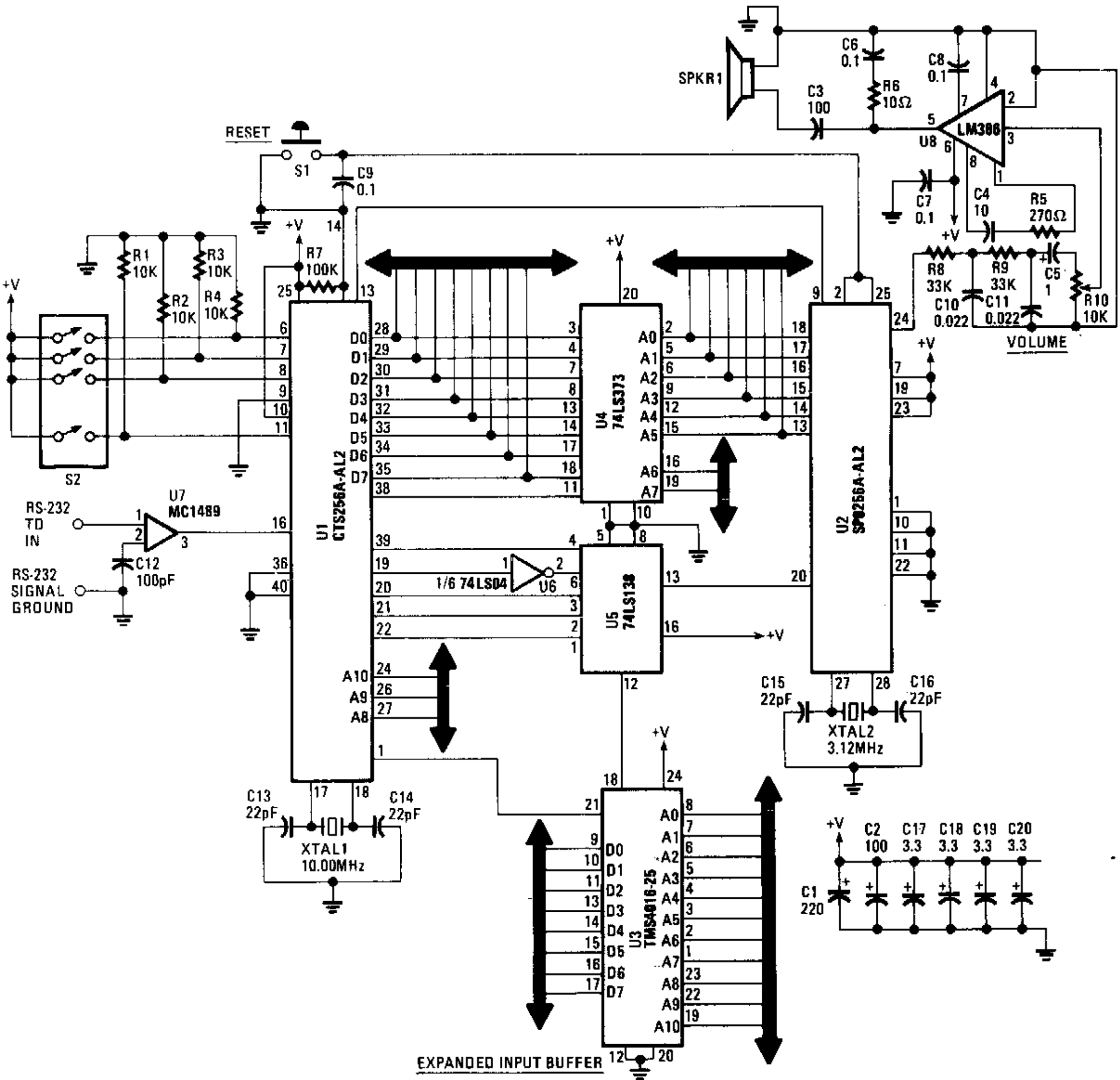


EDN

Fig. 120-2

By vocalizing the numbers and symbols of its keypad, the phone provides an audible confirmation that is useful to the blind. The serial-interface, 2 K-byte × 8-bit ROM (IC4) stores programmed sequences of instructions that are executed by the speech-processor chip IC2—manufactured by the General Instrument Corp. When you depress a key, tone-dialing chip IC1 issues the corresponding number of pulses at its DP output. Counter IC5 totals the pulses, and IC6 latches the resulting 4-bit digital word. This word, converted to serial format by IC2, becomes an address that selects a block of memory within IC4.

COMPUTER SPEECH SYNTHESIZER



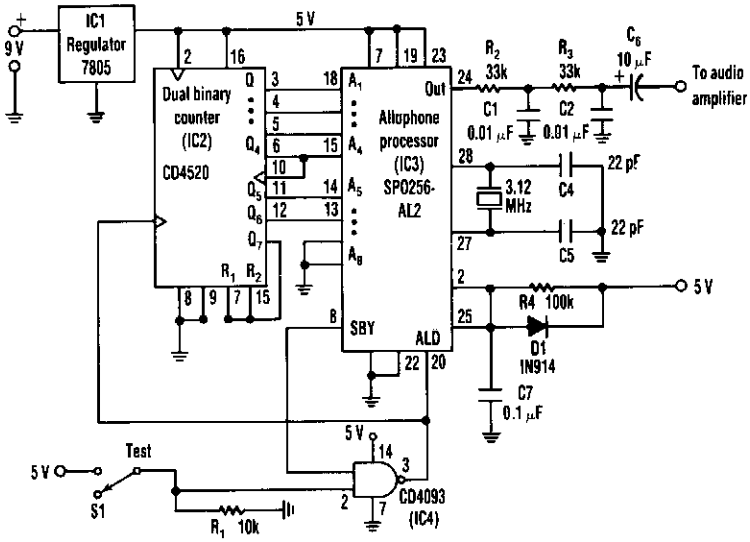
HANDS-ON ELECTRONICS

Fig. 120-3

This text-to-speech converter is built around the SPO256-AL2 speech processor and the CT6256-AL2 text-to-speech converter chips—manufactured by General Instruments. The circuit is set up to receive standard ASCII code from virtually any microcomputer or dumb terminal that is equipped with an RS-232 port—such as a serial-printer or modem port. If a microcomputer is used, the synthesizer can be activated from a terminal-emulator of any communications program, or from any programming language such as BASIC.

The serial input from the RS-232 port enters the circuit through U7, the MC1489 RS-232 receiver chip, and is converted from an RS-232 level to a TTL-level signal. The CTS256-AL2 chip, U1, then converts the ASCII characters into allophone codes and sends those codes to U3, the TMS4016 external-RAM chip. The codes are then transferred to the SPO256-AL2, U2, through the 74LS373 octal latch, U4. Finally, the SPO256-AL2 sends out an audio signal to the LM386 audio amplifier, U8, through some high-pass filtering, and on to the speaker. The 74LS138d, U5, and the 74LSO4, U6, provide control logic.

ALLOPHONE GENERATOR



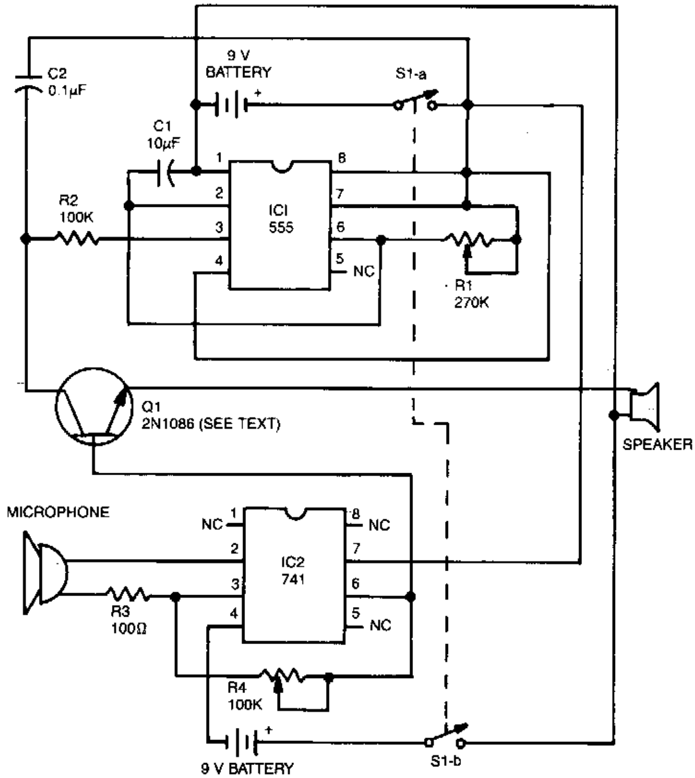
ELECTRONIC DESIGN

Fig. 120-4

The circuit, a general-purpose system with many uses, vocalizes 59 allophones contained in the speech processor. After filtering and amplification, its pulse-code-modulated output can drive an 8-Ω speaker. The processor's address pins, A1 to A6, define 64 speech-entry points.

Closing the test switch to the NAND gate lowers its output, thereby loading an address and triggering the ALD input for an allophone cycle. The CD4520 dual binary counter, IC2, counts from 0 to 63 in binary code until its Q7 output resets it on the number 64 count. To generate a phrase, just add an EPROM between IC2 and IC3 that contains a program for a predetermined sequence of allophones.

ELECTRONIC VOICE SUBSTITUTE



GERNSBACK PUBLICATIONS INC. *Fig. 120-5*

The 555 acts as the tone generator configured in the astable mode. Its pin 3, square-wave output is transformed into a triangle wave by R1 and C2. The *voice's* pitch is controlled by R1. Transistor Q1 can be 2N1086, 2N1091, or any other equivalent npn germanium type. Sounds are amplified by the 741, and the IC's output drives the transistor to saturation. When the transistor is in the saturated state, the triangle wave is able to reach the speaker, and your new *voice* can be heard.

121

Voltage-Controlled Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

BALANCED TMOS VCO

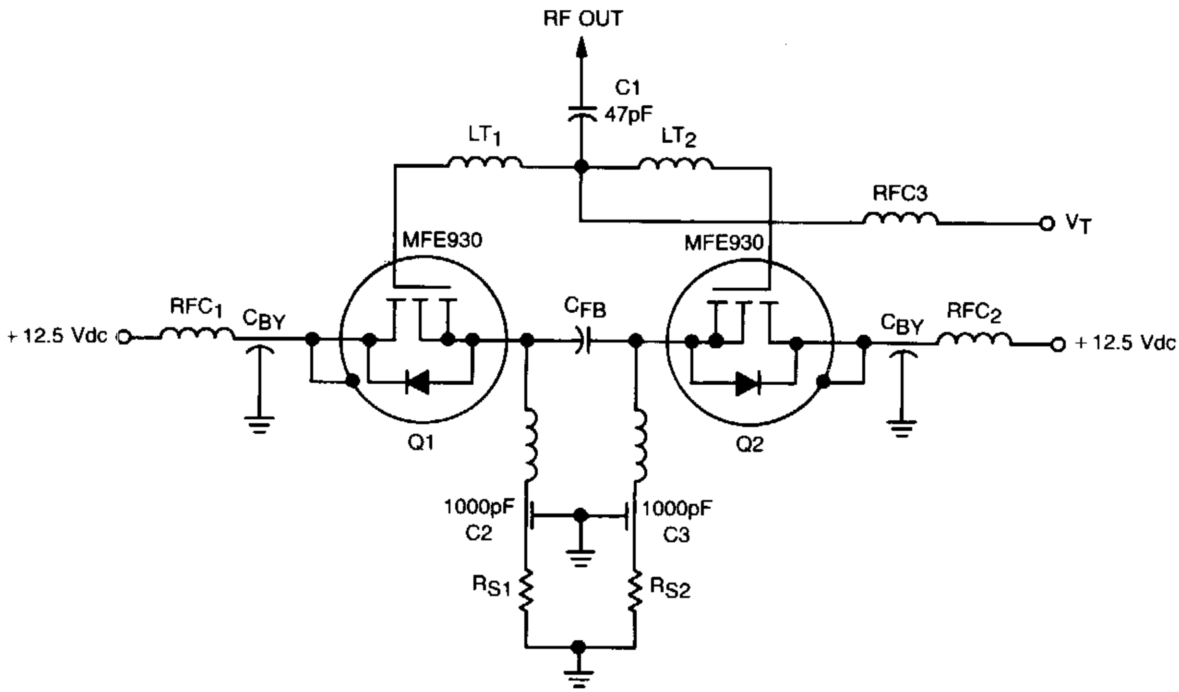


Fig. 121-1

This TMOS VCO operates in push-pull to produce 4 W at 70 MHz. It consists of two MFE930 TMOS devices in a balanced VCO that generally provide better linearity than the single-ended types. Varactors are not used because the design takes advantage of the large change in *Miller* capacitance, C_{RSS}, that is available in TMOS gate structures.

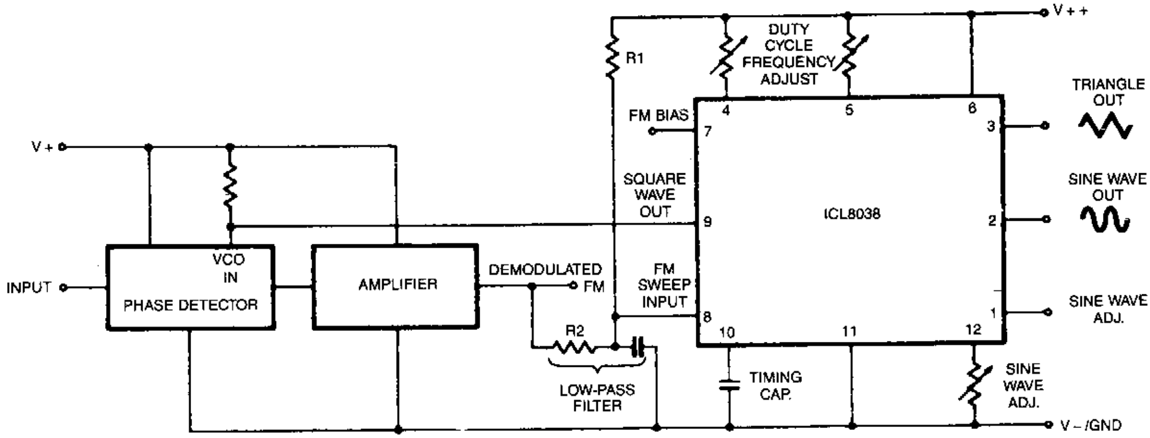
In the balanced VCO, the fundamental (f_O) and/or twice the fundamental ($2f_O$) can be coupled from the circuit at separate nodes. This makes the balanced oscillator very useful in phase-locked loops. The fundamental:

$$f_O = \frac{1}{2} (L_F C_{RSS})^{-1/2}$$

where:

$$L_F = 0.68\ \mu\text{H}$$

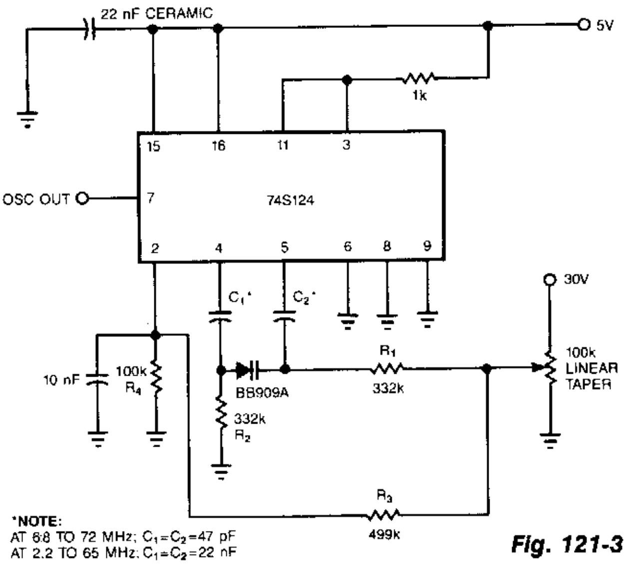
WAVEFORM GENERATOR/STABLE VCO



INTERSIL *Fig. 121-2*

In this circuit, a waveform generator is used as a stable VCO in a Phase-Locked Loop (PLL).

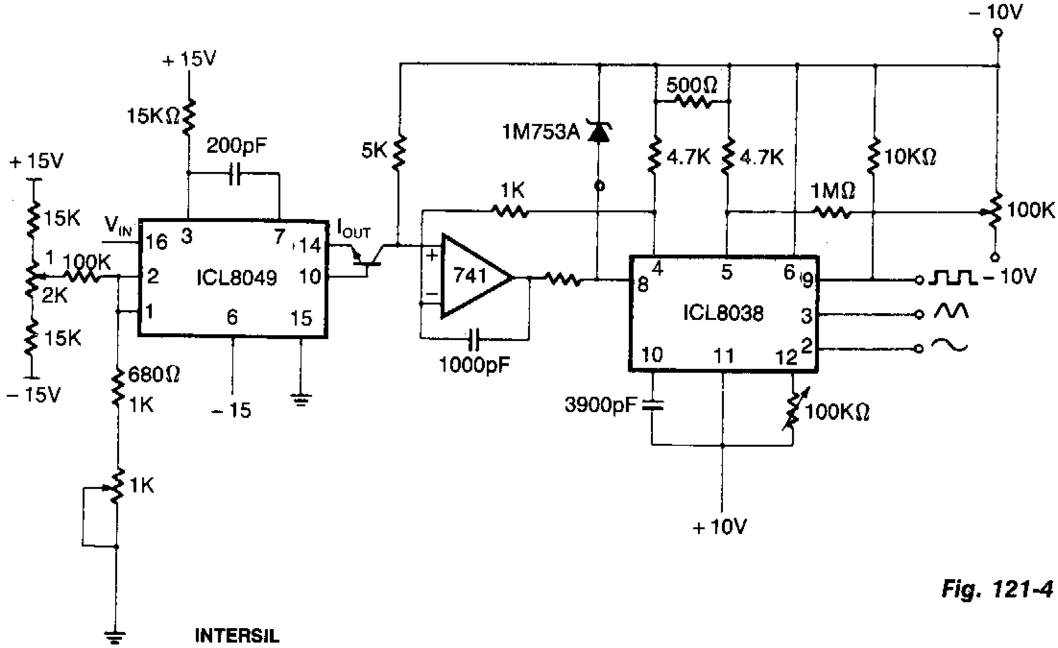
VARIABLE-CAPACITANCE DIODE-SPARKED VCO



EDN *Fig. 121-3*

You can transform a 741S124 multivibrator into a wideband VCO by replacing it conventional fixed capacitor with a variable-capacitance diode. The only disadvantage of this scheme is the 30-V biasing voltage that the diode requires. Capacitors C1 and C2 couple the Philips BB909A variable-capacitance diode to the 74S124. R1 and R2 are large enough to isolate ground and control voltages from the timing capacitors. Resistors R3 and R4 form a voltage divider for the 74S124's control input.

LOGARITHMIC SWEEP VCO

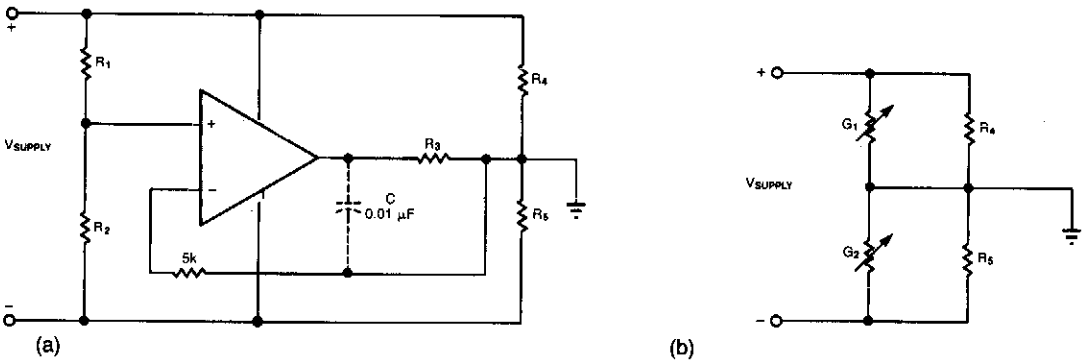


INTERSIL

Fig. 121-4

This circuit uses the output of the ICL8049 to control the frequency of the ICL3038 waveform generator; the 741 op amp is used to linearize the voltage-frequency response. The input voltage to the 8049 can be, for example, from the horizontal sweep signal of an oscilloscope; the output of the 8038 will then sweep logarithmically across the audio range. By feeding this to the equipment being measured and detecting the output, a standard frequency response can be obtained. If the output is fed through an ICL8048 before being displayed, a standard bode plot results.

SUPPLY VOLTAGE SPLITTER

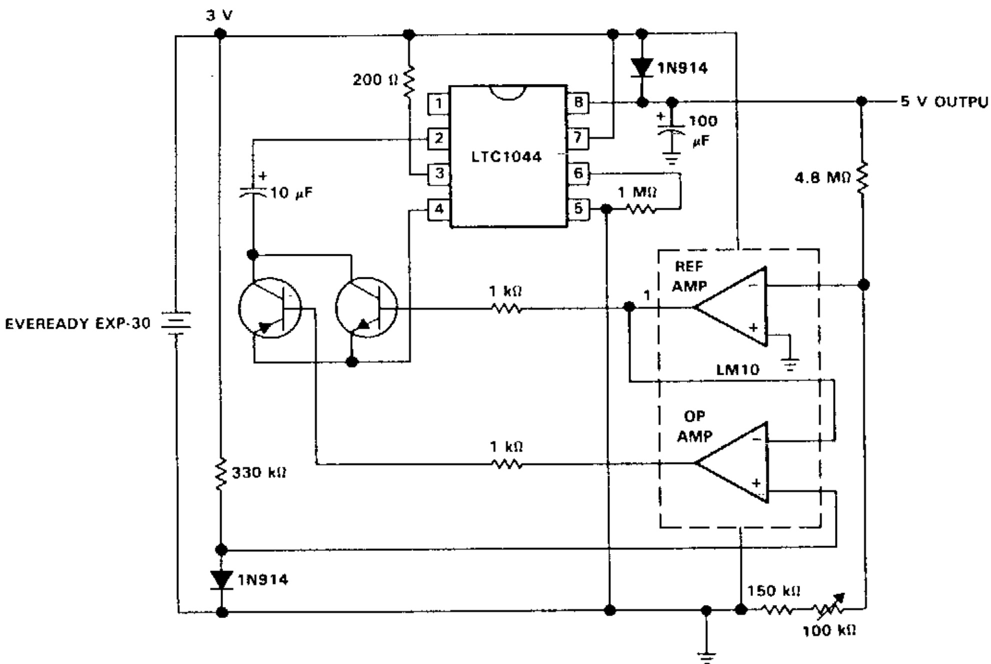


EDN

Fig. 121-5

This simple circuit can convert a single supply voltage, such as a battery, into a bipolar supply. Sense resistors R1 and R2 establish relative magnitudes for the resulting positive and negative voltages. Their rail-to-rail value, of course, equals V_{SUPPLY}. R4 and R5 represent the load impedances. For example, equal-value sense resistors produce 1/2 V_{SUPPLY} across each of the load resistors, R4 and R5. The op amp maintains these equal voltages by sinking or sourcing current through R3; the op amp's action is equivalent to that of variable conductances G1 and G2 in shunt with each load resistor. Choose a value for R3 so that the largest voltage across it, the greatest load-current mismatch, won't exceed the op amp's output-voltage capability for the application. You can add a buffer amplifier at the op amp's output to provide greater load currents. If you need bypass capacitors across the load resistors as well, connect a capacitor (dashed lines) to ensure that the amplifier remains stable.

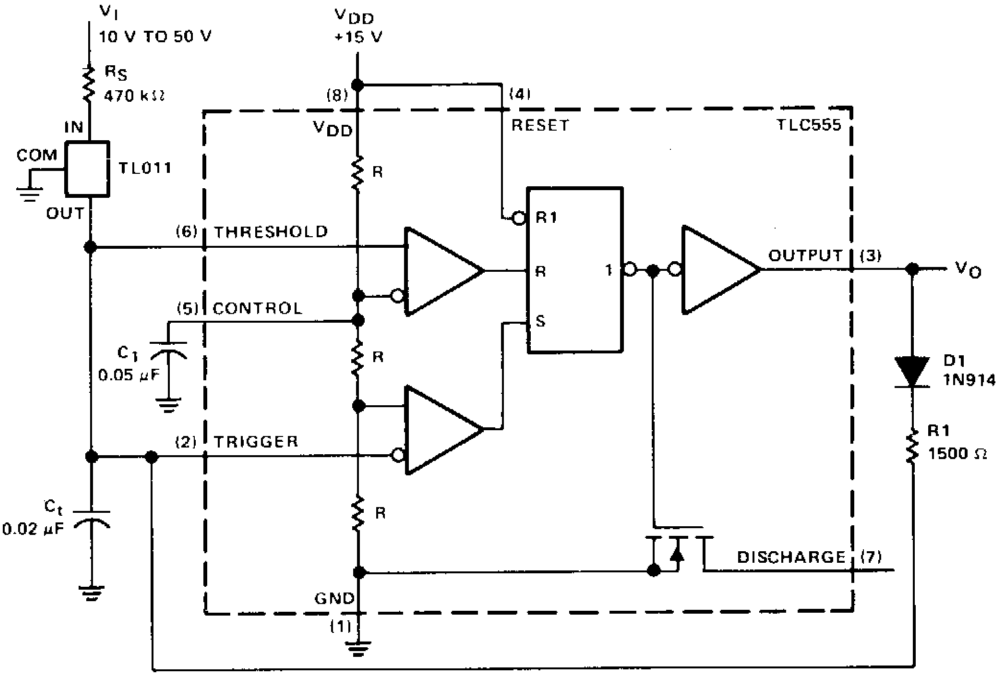
3 – 5 V REGULATED OUTPUT CONVERTER

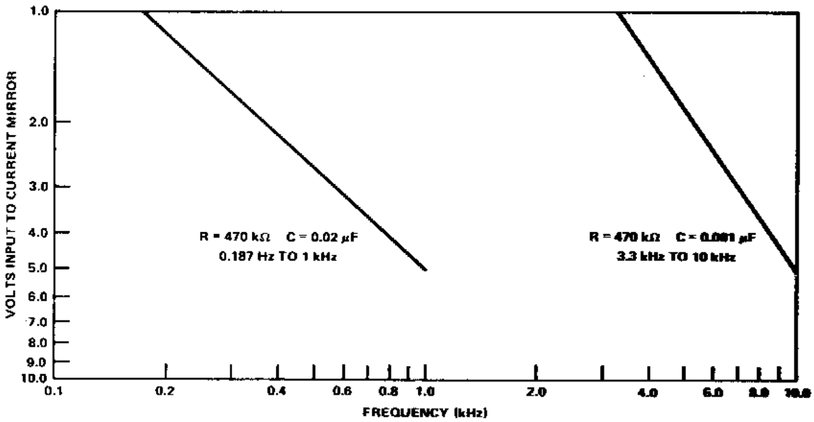


Reprinted by permission of Texas Instruments.

Fig. 121-6

VCO





Voltage-Controlled Oscillator Frequency vs Voltage

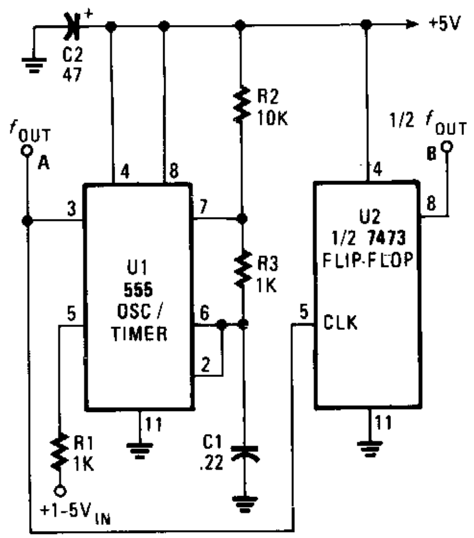
Reprinted by permission of Texas Instruments.

Fig. 121-7

At startup, the voltage in the trigger input at pin 2 is less than the trigger level voltage, 1/3 V_{DD}, causing the timer to be triggered via pin 2. The output of the timer at pin 3 becomes high, allowing capacitor C_t to charge very rapidly through diode D1 and resistor R1.

When capacitor C_t charges to the upper threshold voltage 2/3 V_{DD}, the flip-flop is reset, the output at pin 3 decreases, and capacitor C_t discharges through the current mirror, TL011. When the voltage at pin 2 reaches 1/3 V_{DD}, the lower threshold or trigger level, the timer triggers again and the cycle is repeated.

SIMPLE VCO



POPULAR ELECTRONICS

Fig. 121-8

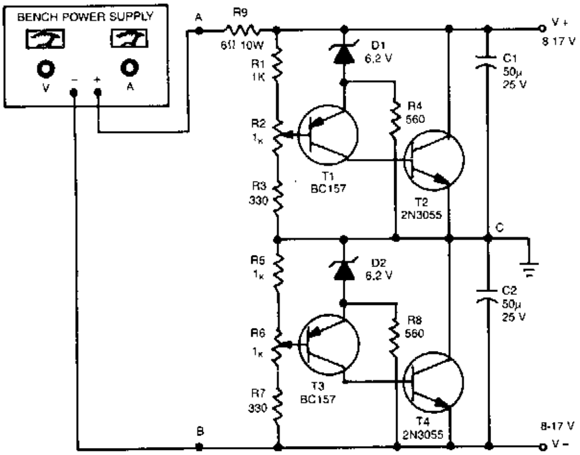
The output frequency of the VCO, U1, varies inversely with the input voltage. With a 1-V input, the oscillator output frequency is about 1500 Hz; with a 5-V input, the output frequency drops to around 300 Hz. The output frequency range of U1 can be altered by varying the values of C1, R2, and R3. Increasing the value of any those three components will lower the oscillator frequency, and decreasing any of those values will raise the frequency. Output-waveform symmetry suffers since the frequency varies from one extreme to the other. At the highest frequency, the waveform is almost equally divided. But when the frequency drops, the output of the circuit turns into a narrow pulse. If a symmetrical waveform is required, add the second IC, U2, half of a 7473P dual TTL J-K flip-flop, to the oscillator circuit. The signal frequency output by U2 is 1/2 of the input.

122

Voltage Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

UNIPOLAR-TO-DUAL SUPPLY CONVERTER

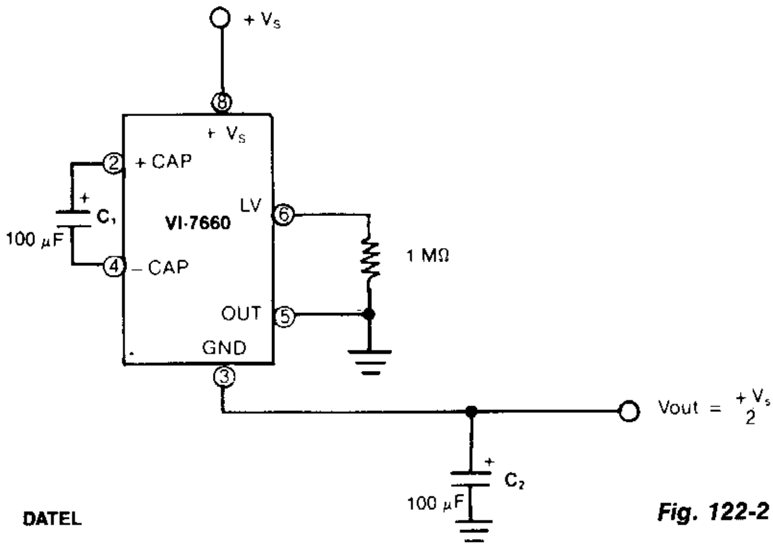


ELECTRONIC ENGINEERING

Fig. 122-1

The outputs in this circuit are independently variable and can be loaded unsymmetrically. The output voltage remains constant, irrespective of load and changes. By varying potentiometers R2 or R6, the output voltages can be conveniently set. Outputs can be varied between 8 and 17 V, so that the standard ±9, ±12, and ±15 V settings can be made. This converter is designed for a maximum load current of 1 A and the output impedance of both supplies of 0.35 Ω. This circuit is not protected against shortcircuits, but uses the protection provided by the dc input source. This circuit is ideal for biassing operation amplifier circuits.

EFFICIENT SUPPLY SPLITTER



DATEL

Fig. 122-2

In this application, the VI-7660 is connected as a voltage splitter. Note that the *normal* output pin is connected to ground and the *normal* ground pin is used as the output. The switches that allow the charge pumping are bidirectional; therefore, charge transfer can be performed in reverse. The 1-MΩ resistor is used to avoid start-up problems by forcing the internal regulator on. An application for this circuit would be driving low-voltage, ±7.5 Vdc, circuits from ±15 Vdc supplies, or low-voltage logic from 9 to 12 V batteries.

HIGH-EFFICIENCY FLYBACK VOLTAGE CONVERTER

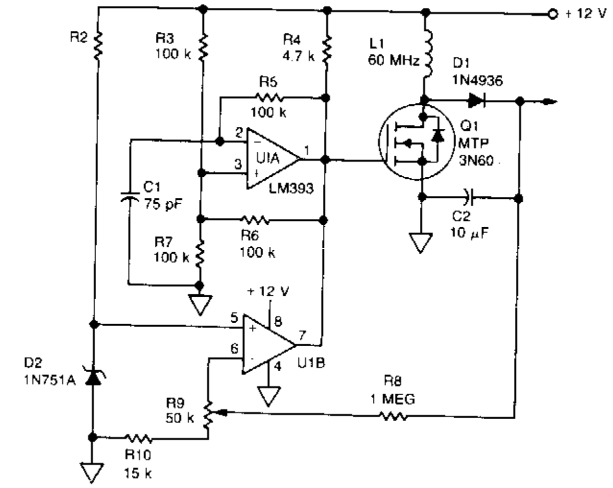


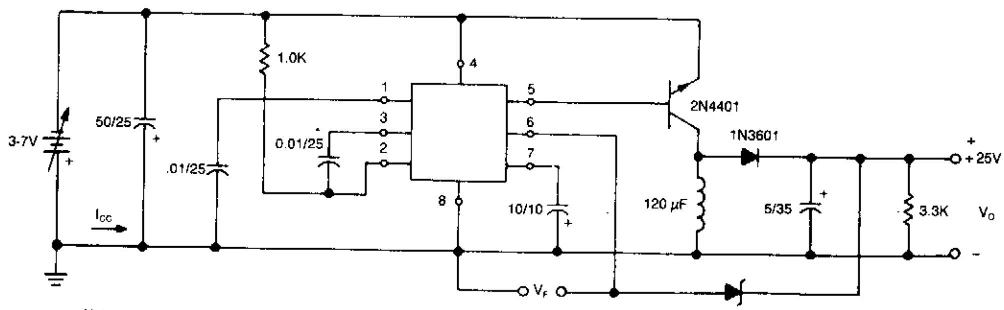
Fig. 122-3

U1 is a dual voltage comparator with open collector outputs. The A side is an oscillator operating at 100 kHz, and the B side is part of the regulation circuit that compares a fraction of the output voltage to a reference generated by zener diode D2.

The output of U1A is applied directly to the gate of Q1. During the positive half-cycle of the Q1 gate voltage, energy is stored in L1; in the negative half, the energy is discharged into C2. A portion of the output voltage is fed back to U1B to provide regulation. The output voltage is adjustable by changing feedback potentiometer R9.

Using the component values shown will produce a nominal 300-V output from a 12-V source. However, the circuit maximum output voltage is limited by R10; a lower value for R10 yields a higher output voltage. The output voltage is also limited by the breakdown of values Q1, L1, D1, and C2.

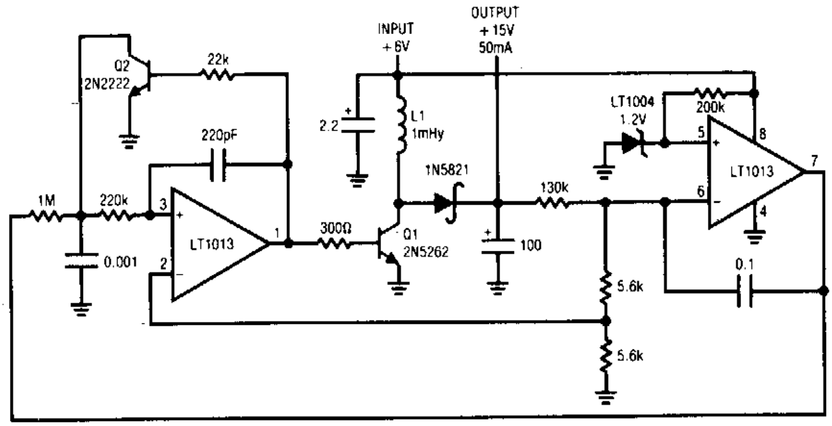
3 – 25 V DC-DC CONVERTER



Notes:
1. All resistor values in ohms, ± 1%, 1/4W
2. All capacitor values in μF, ± 20%, except * ± 5%.
3. All inductors ± 4%.
4. IC is MC3380P

Fig. 122-4

REGULATED 15-V_{OUT} 6-V DRIVEN CONVERTER



LINEAR TECHNOLOGY CORP.

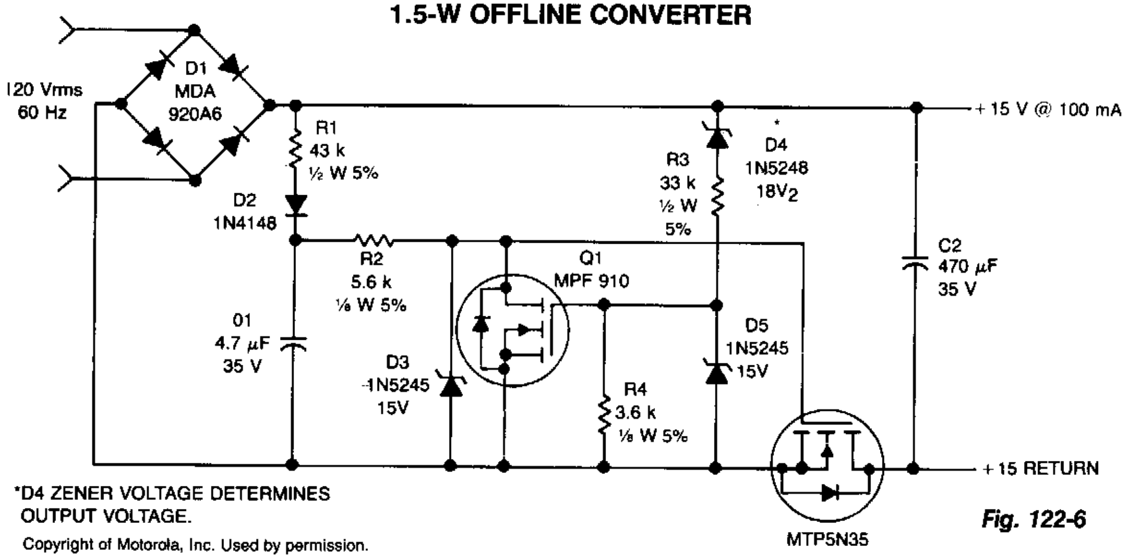
Fig. 122-5

This converter delivers up to 50 mA from a 6-V battery with 78% efficiency. This flyback converter functions by feedback-controlling the frequency of inductive flyback events. The inductor's output, rectified and filtered to dc, biases the feedback loop to establish a stable output. If the converter's output is below the loop setpoint, A2's inputs unbalance and current is fed through the 1-MΩ resistor at A1. This ramps the 1000-pF value positive. When this ramp exceeds the 0.5-V potential at A1's positive input, the amplifier switches high. Q2 turns on, discharging the capacitor to ground. Simultaneously, regenerative feedback through the 200-pF value causes a positive-going pulse at A1's positive input, sustaining A1's positive output. Q1 comes on, allowing inductor, L1, current to flow. When A1's feedback pulse decays, its output becomes low, turning off Q1. Q1's collector is pulled high by the inductor's flyback and the energy is stored in the 100-μF capacitor. The capacitor's voltage, which is the circuit output, is sampled by A2 to close a loop around A1/Q1. This loop forces A1 to oscillate at whatever frequency is required to maintain the 15-V output.

In-phase transformer windings for the drain and gate of TMOS power FET Q1 cause the circuit to oscillate. Oscillation starts when the feedback coupling capacitor, C1, is charged from the supply line via a large resistance; R2 and R3 limit the collector current to Q2. During *pump-up*, the on time is terminated by Q2, which senses the ramped source current of Q1. C1 is charged on alternate half-cycles by Q2 and forward-biased by zener D2.

When the regulated level is reached, forward bias is applied to Q2, terminating the on time earlier at a lower peak current. When this occurs, the frequency increases in inverse proportion to current, but the energy per cycle decreases in proportion to current squared. Therefore, the total power coupled through the transformer to the secondary is decreased.

1.5-W OFFLINE CONVERTER



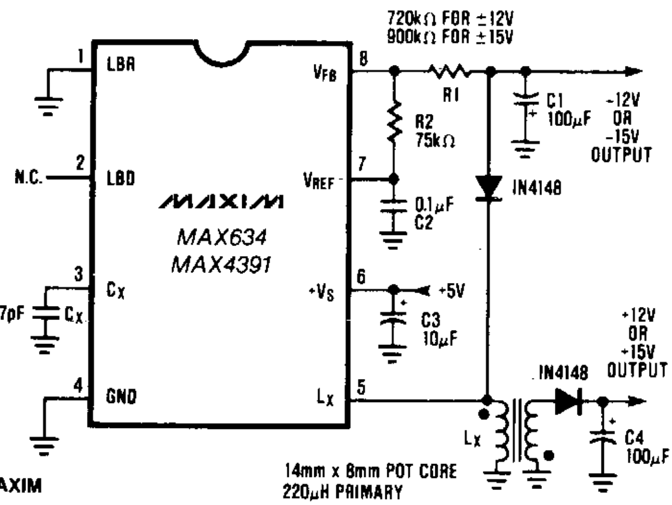
*D4 ZENER VOLTAGE DETERMINES OUTPUT VOLTAGE.

Fig. 122-6

This nonisolated, unregulated, minimum component converter fills the void between low-power zener regulation and the higher power use of a 60-Hz input transformer. It is intended for use wherever a nonisolated supply can be used safely.

The circuit operates by conducting only during the low-voltage portion of the rectified sine wave. R1 and D2 charge C1 to approximately 20 V, which is maintained by Q1. This voltage is applied to the gate of Q2, turning it on. When the rectified output voltage exceeds the zener voltage of D4, Q1 turns on, shunting the gate of Q2 to ground, turning it off.

DUAL OUTPUT ±12 OR ±15 V DC-DC CONVERTER

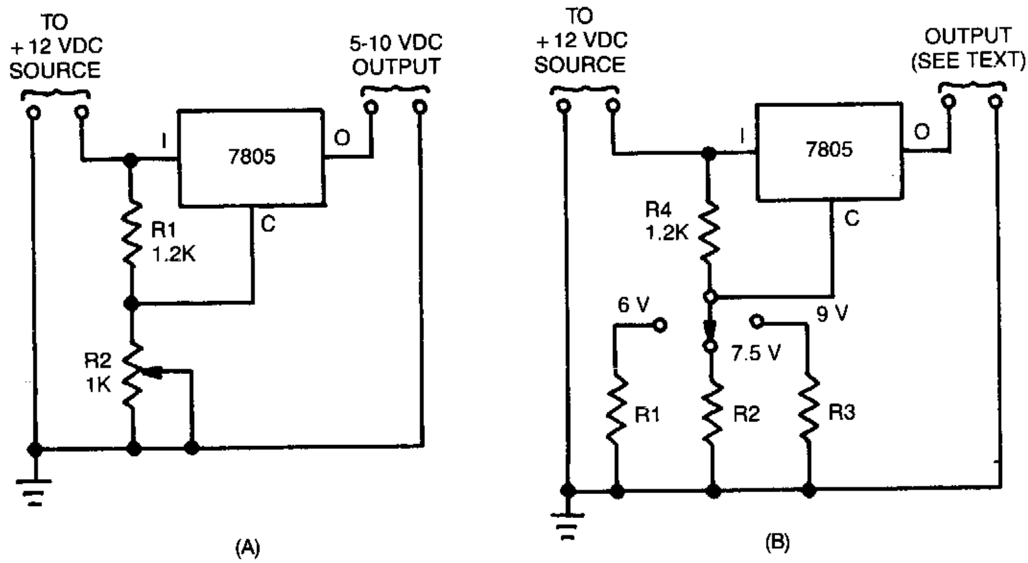


MAXIM

Fig. 122-7

The buck-boost configuration of the MAX634 is well suited for dual output dc-dc converters. Only a second winding on the inductor is needed. Typically, this second winding is bifilar—primary and secondary are wound simultaneously using two wires in parallel. The inductor core is usually a toroid or a pot core. The negative output voltage is fully regulated by the MAX634. The positive voltage is semiregulated, and will vary slightly with load changes on either the positive or negative outputs.

12-TO-16 V CONVERTER



Reprinted with permission of Radio-Electronics Magazine, April 1985. Copyright Gernsback Publications, Inc., 1985. *Fig. 122-8*

Many devices operate from a car's 12-V electrical system. Some require 12 V; others require some lesser voltage. An automobile battery's output can vary from 12 to 13.8 V under normal circumstances. The load requirements of the device might vary. This circuit maintains a constant voltage regardless of how those factors change. Simple circuit, A, uses a 7805 voltage regulator. In addition to a constant output, this IC provides overload and short-circuit protection. That unit is a 5-V, 1-A regulator, but when placed in circuit B, it can provide other voltages as well. When the arm of potentiometer R1 is moved toward ground, the output varies from 5 to about 10 V.

SELF-OSCILLATING FLYBACK-SWITCHING CONVERTER

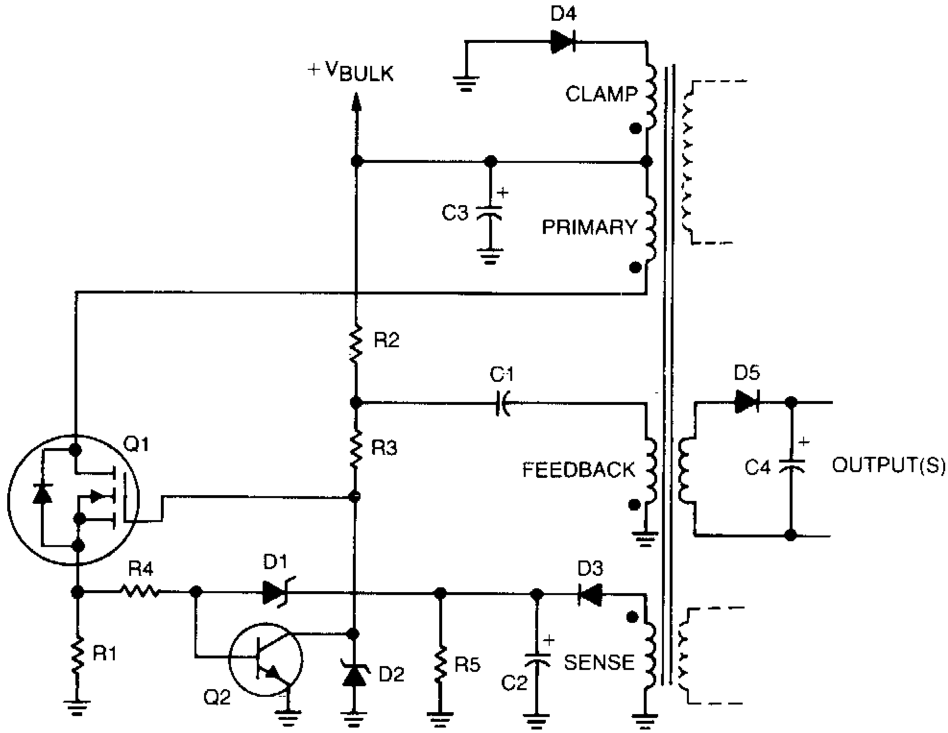


Fig. 122-9

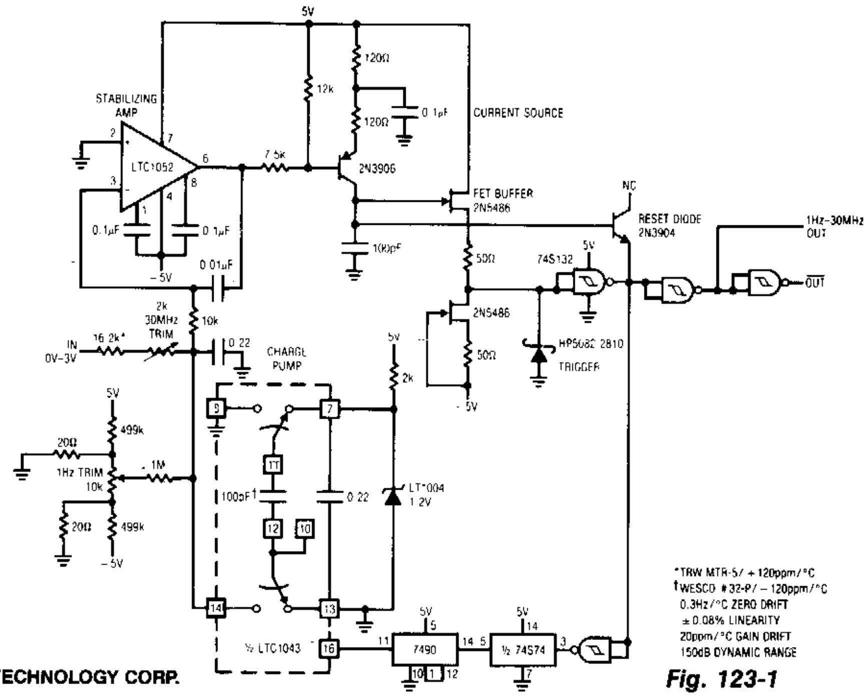
Regulation is provided by taking the rectified output of the sense winding and applying it as a bias to the base of Q2 via zener D1. The collector of Q2 then removes drive from the gate of Q1. Therefore, if the output voltage should increase, Q2 removes the drive to Q1 earlier, shortening the on time, and the output voltage will remain the same. Dc outputs are obtained by merely rectifying and filtering secondary windings, as done by D5 and C4.

123

Voltage-to-Frequency Converters

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

1 Hz-TO-30 MHz VOLTAGE-TO-FREQUENCY CONVERTER

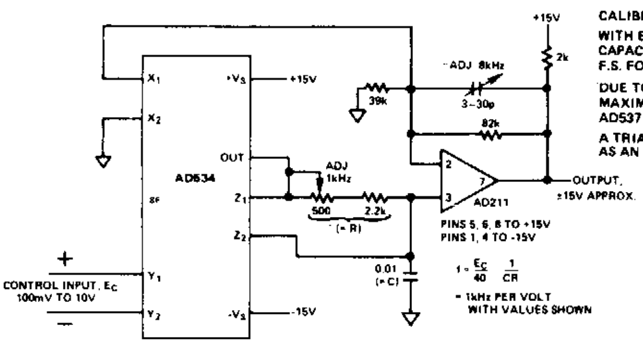


LINEAR TECHNOLOGY CORP.

Fig. 123-1

Circuit has a 1 Hz-to-30 MHz output, 150-dB dynamic range, for a 0 to 5 V input. It maintains 0.08% linearity over its entire 7¹/₃ decade range with a full-scale drift of about 20 ppm/°C. To get the additional bandwidth, the fast JFET buffer drives the Schottky TTL Schmitt trigger. The Schottky diode prevents the Schmitt trigger from ever seeing negative voltage at its input. The Schmitt's input voltage hysteresis provides the limits which the oscillator runs between. The 30-MHz, full-scale output is much faster than the LTC1043 can accept, so the digital divider stages are used to reduce the feedback frequency signal by a factor of 20. The remaining Schmitt sections furnish complementary outputs.

DIFFERENTIAL-INPUT VOLTAGE-TO-FREQUENCY CONVERTER



CALIBRATION PROCEDURE:

WITH E_C = 1.0V, ADJUST POT TO SET f = 1.000kHz. WITH E_C = 8.0V, ADJUST TRIMMER CAPACITOR TO SET f = 8.000kHz. LINEARITY WILL TYPICALLY BE WITHIN ±0.1% OF F.S. FOR ANY OTHER INPUT.

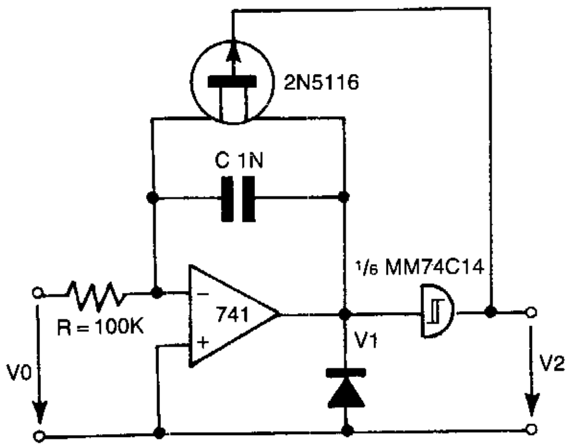
DUE TO DELAYS IN THE COMPARATOR, THIS TECHNIQUE IS NOT SUITABLE FOR MAXIMUM FREQUENCIES ABOVE 10kHz. FOR FREQUENCIES ABOVE 10kHz THE AD537 VOLTAGE TO FREQUENCY CONVERTER IS RECOMMENDED.

A TRIANGLE-WAVE OF ±5V PK APPEARS ACROSS THE 0.01µF CAPACITOR; IF USED AS AN OUTPUT, A VOLTAGE-FOLLOWER SHOULD BE INTERPOSED.

Reprinted with permission of Analog Devices, Inc.

Fig. 123-2

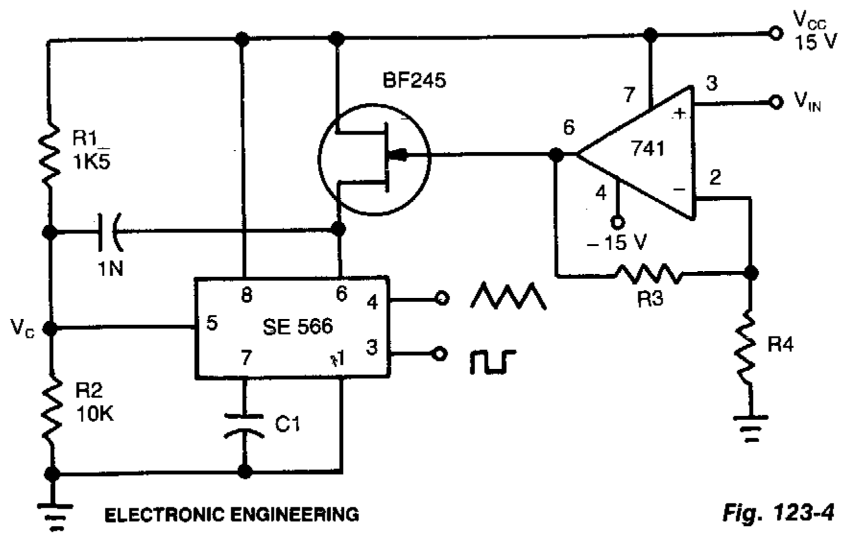
LOW-COST VOLTAGE-TO-FREQUENCY CONVERTER



ELECTRONIC ENGINEERING *Fig. 123-3*

The 741 op amp integrator signal is fed into the Schmitt trigger input of an inverter. When the signal reaches the magnitude of the positive-going threshold voltage, the output of the inverter is switched to zero. The inverter output controls the FET switch directly. For a gate voltage of zero, the FET channel turns on to low resistance and the capacitor is discharged. The discharge current depends on the on resistance of the FET. When the capacitor C1 is discharged to the negative-going threshold voltage level of the inverter, the inverter output is switched to ±12 V. This switch causes the FET channel to be switched off, and the discharging process is switched into a charging process again. Using the components shown, an output frequency of about 10 kHz with 0.1% linearity can be obtained.

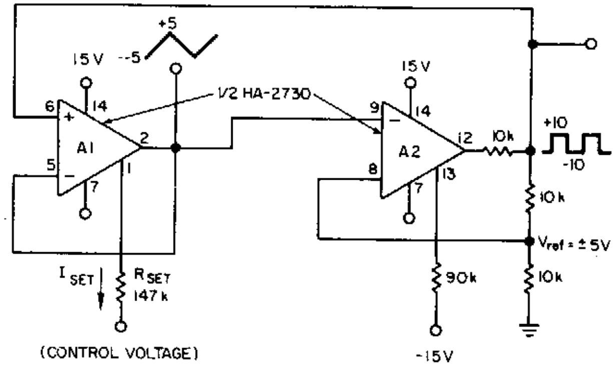
WIDE-RANGE VOLTAGE-TO-FREQUENCY CONVERTER



ELECTRONIC ENGINEERING *Fig. 123-4*

This circuit is based upon the change of frequency of the function generator with the input voltage V_{IN}. Generally, the frequency depends upon the capacitance and resistor connected to pin 6. This resistor is replaced by the FET. The frequency range is adjustable by changing the input voltage, V_{IN}; the converter will give a range of 10 Hz to 1 MHz.

WIDE-RANGE VOLTAGE-TO-FREQUENCY CONVERTER

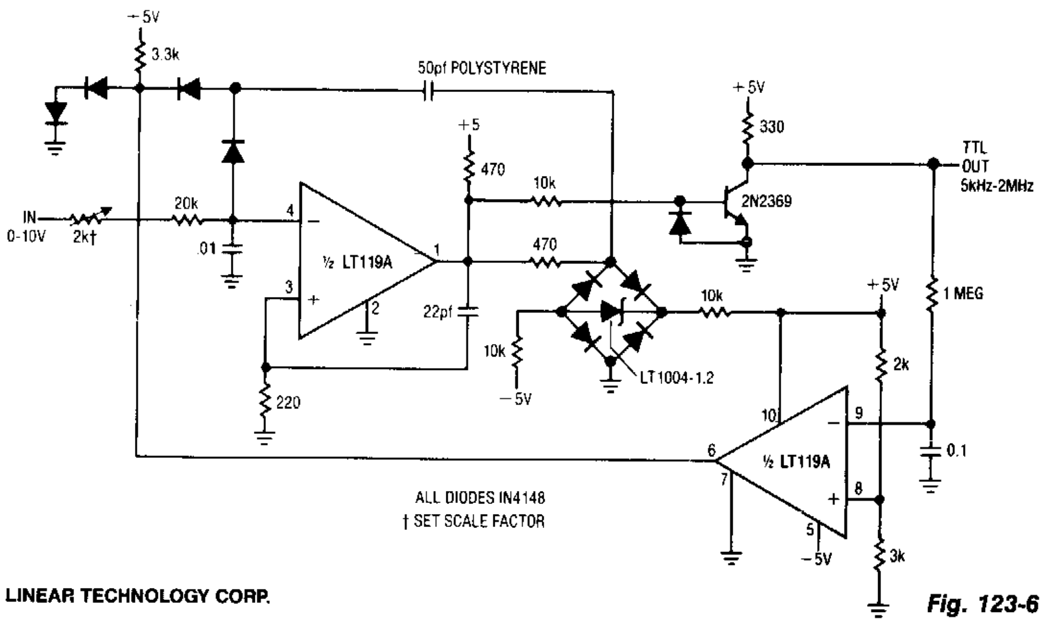


ELECTRONIC DESIGN

Fig. 123-5

This circuit uses a programmable op amp such as the HA2730—a two-amplifier monolithic chip with independent programming ports for each amplifier—whose slew rate and other parameters vary linearly with a so-called *set current*. The converter circuit uses one amplifier, A1, as a slewing amplifier and other, A2, as a comparator function. The control voltage V_C, determines A1's slew rate. And, because A1's output voltage swing is constant, the modulation of its set current results in direct control of the circuit's frequency. A1's internal compensation capacitor acts as the timing component. An internal bipolar current source, whose current magnitude is directly proportional to the set current of pin 1, then determines the charge-discharge rate. A conversion nonlinearity of $\pm 0.03\%$ of full scale over 3 decades and $\pm 1.5\%$ of full scale over 4.3 decades of frequency is possible. The frequency range is adjustable by a change in the resistance, R.

5 KHz-TO-2 MHz VOLTAGE-TO-FREQUENCY CONVERTER



LINEAR TECHNOLOGY CORP.

Fig. 123-6

PRESERVED INPUT VOLTAGE-TO-FREQUENCY CONVERTER

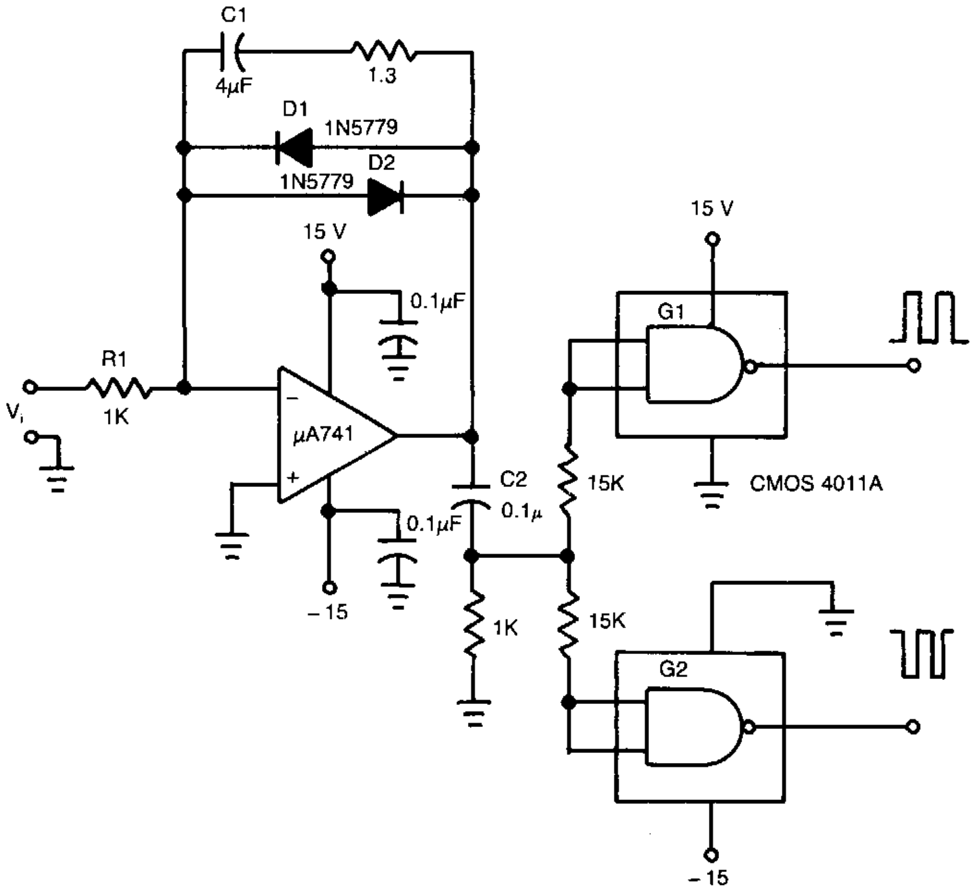


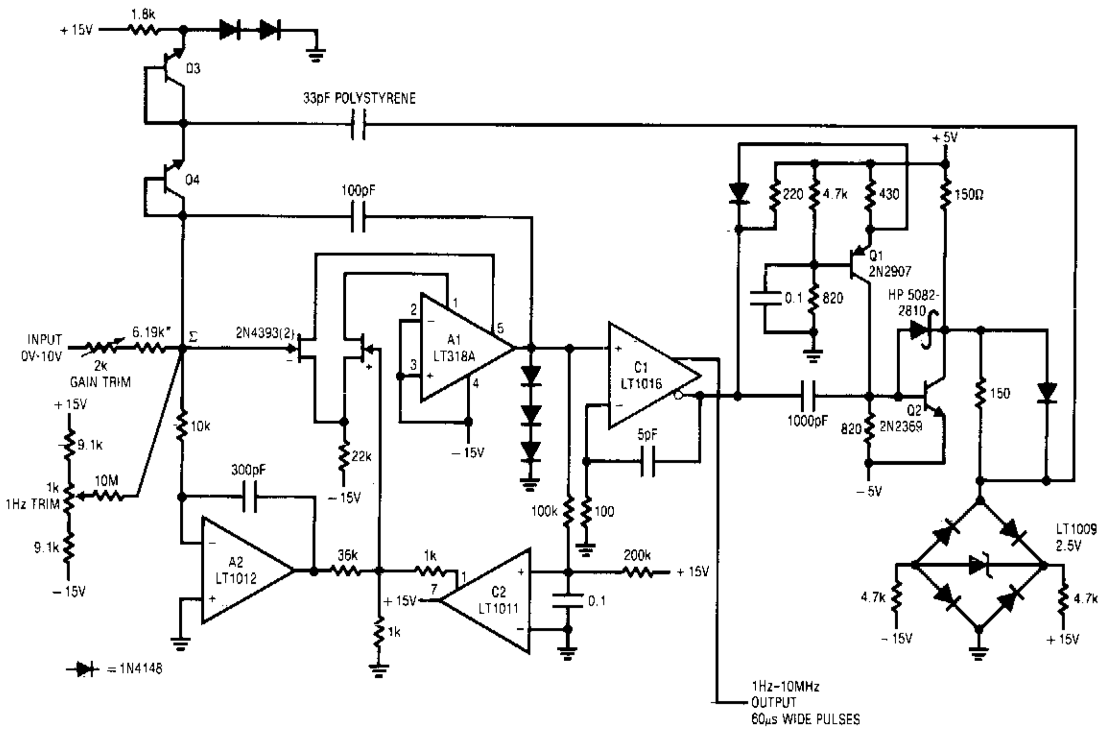
Fig. 123-7

ELECTRONIC DESIGN

The input voltage, V1, causes C1 to charge and produce a ramp voltage at the output of the 741 op amp. Diodes D1 and D2 are four-layer devices. When the voltage across C1 reaches the breakover voltage of either diode, the diode conducts to discharge C1 rapidly and the op amp output goes abruptly to zero. This rapid discharge action applies a narrow pulse to G1 and G2. Positive discharge pulses produced by a positive V1 are coupled to the output only through G1, while negative pulses are coupled only through G2.

Because of the forward breakover current of diodes D1 and D2, the circuit won't operate below a minimum input voltage. An increase of R1 increases this minimum voltage and reduces the circuit's dynamic range. The minimum input voltage with R1 at 1 KΩ is in the range of 10 to 50 mV. This input dead zone, when input signal V1 is near zero is desirable in applications that require a signal to exceed a certain level before an output is generated.

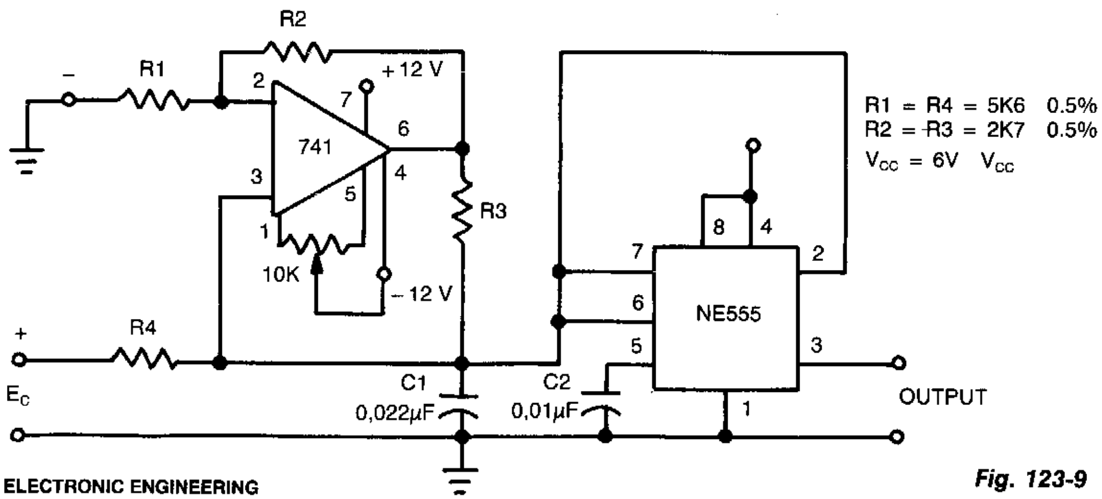
1 Hz-TO-10 MHz VOLTAGE-TO FREQUENCY CONVERTER



LINEAR TECHNOLOGY CORP.

Fig. 123-8

VOLTAGE-TO-FREQUENCY CONVERTER



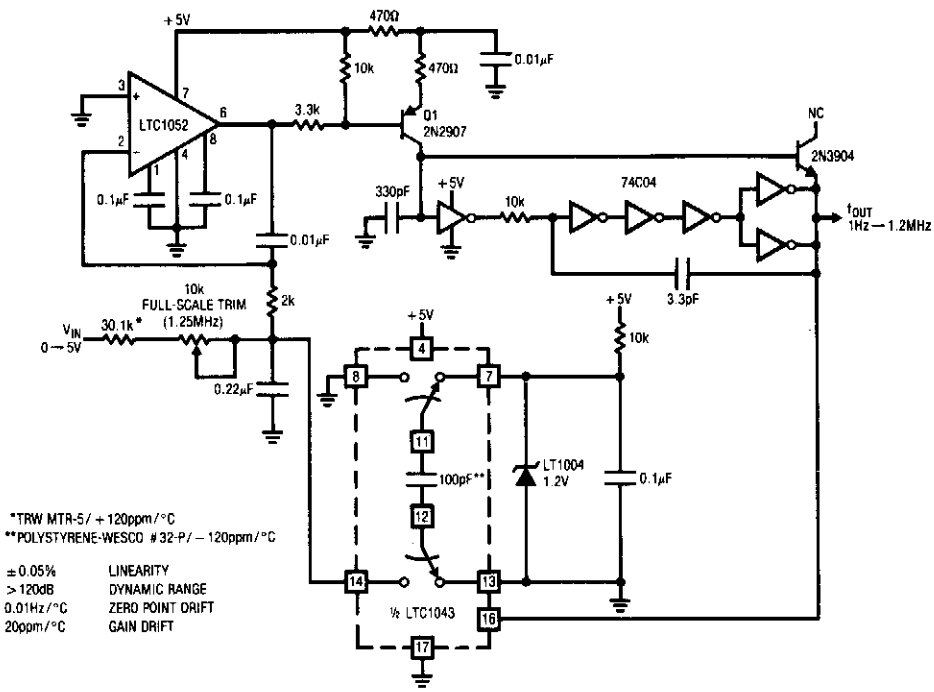
ELECTRONIC ENGINEERING

Fig. 123-9

This circuit can accept positive or negative or differential control voltages. The output frequency is zero when the control voltage is zero. The 741 op amp forms a current source controlled by the voltage E_C to charge the timing capacitor C1 linearly. NE555 is connected in the astable mode, so that the capacitor charges and discharges between $1/3\ V_{CC}$ and $2/3\ V_{CC}$. The offset is adjusted by the 10-K potentiometer so that the frequency is zero when the input is zero. For the component values shown: $f \approx 4.2\ E_C$ kHz. If two dc voltages are applied to the ends of R1 and R4, the output frequency will be proportional to the difference between the two voltages.

---

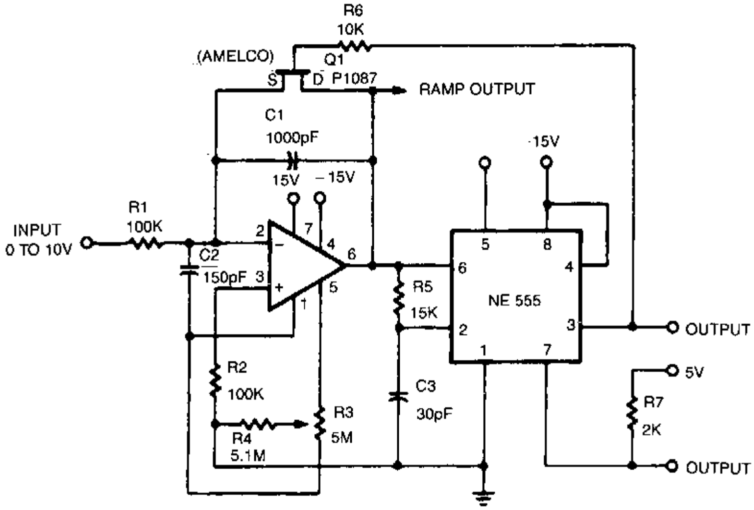
1 HZ-TO-1.25 MHZ VOLTAGE-TO-FREQUENCY CONVERTER



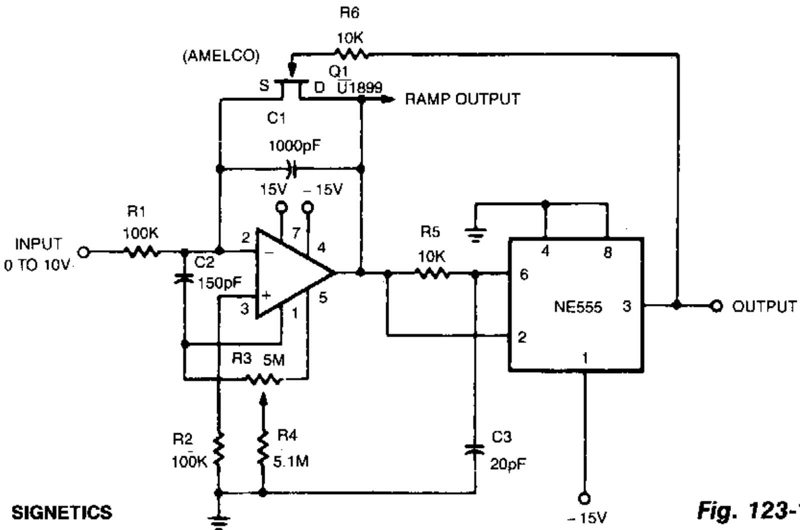
LINEAR TECHNOLOGY CORP. *Fig. 123-10*

This stabilized voltage-to-frequency converter features 1 Hz – 1.25 MHz operation, 0.05% linearity, and a temperature coefficient of typically 20 ppm/°C. This circuit runs from a single 5-V supply. The converter uses a charge feedback scheme to allow the LTC1052 to close a loop around the entire circuit, instead of simply controlling the offset. This approach enhances linearity and stability, but introduces the loop's settling time into the overall voltage-to-frequency step-response characteristic.

ACCURATE VOLTAGE-TO-FREQUENCY CONVERTER



(a)



SIGNETICS

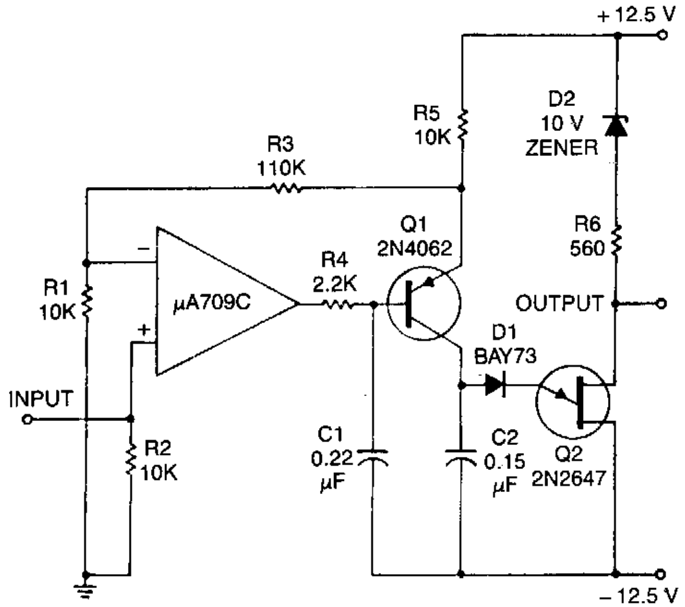
Fig. 123-11

(b)

NOTE:
ALL RESISTOR VALUES IN OHMS

This linear voltage-to-frequency converter, a, achieves good linearity over 0 to −10 V. Its mirror image, b, provides the same linearity over 0 to +10 V, but it is not DTL/TTL compatible.

VOLTAGE-TO-FREQUENCY CONVERTER



ELECTRONIC DESIGN

Fig. 123-12

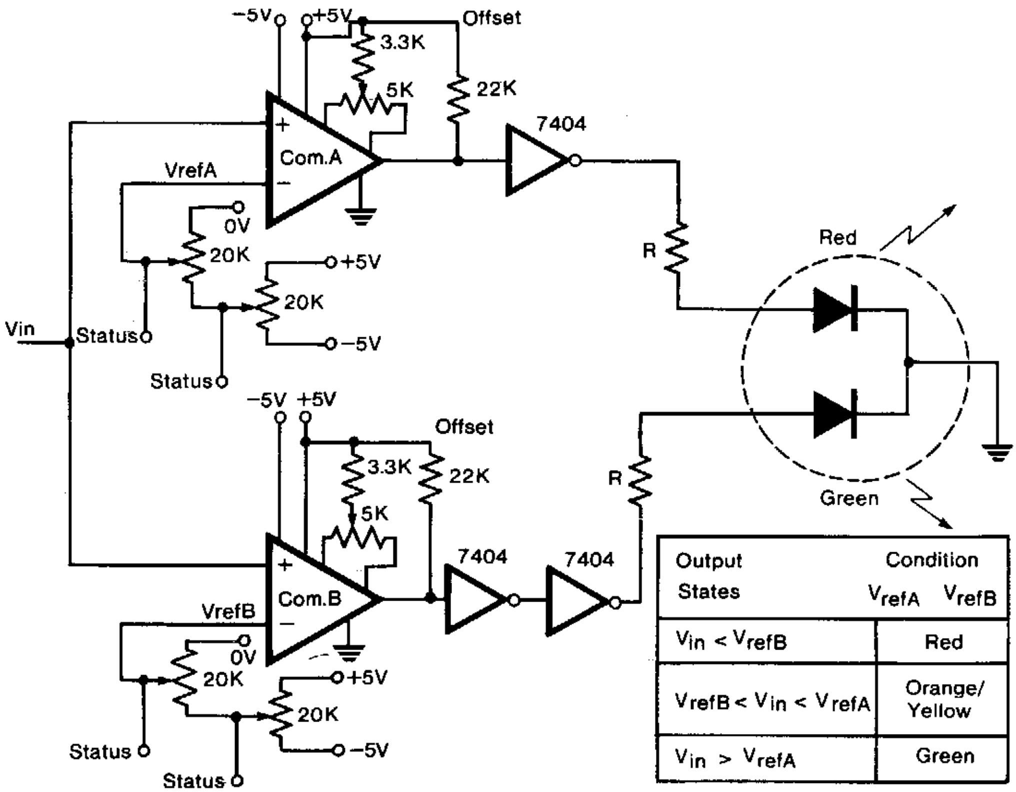
This circuit consists of a UJT oscillator in which the timing charge capacitor C2 is linearly dependent on the input signal voltage. The charging current is set by the voltage across resistor R5, which is accurately controlled by the amplifier.

124

Voltage Meters/Monitors/Indicators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

VOLTAGE-LEVEL INDICATOR



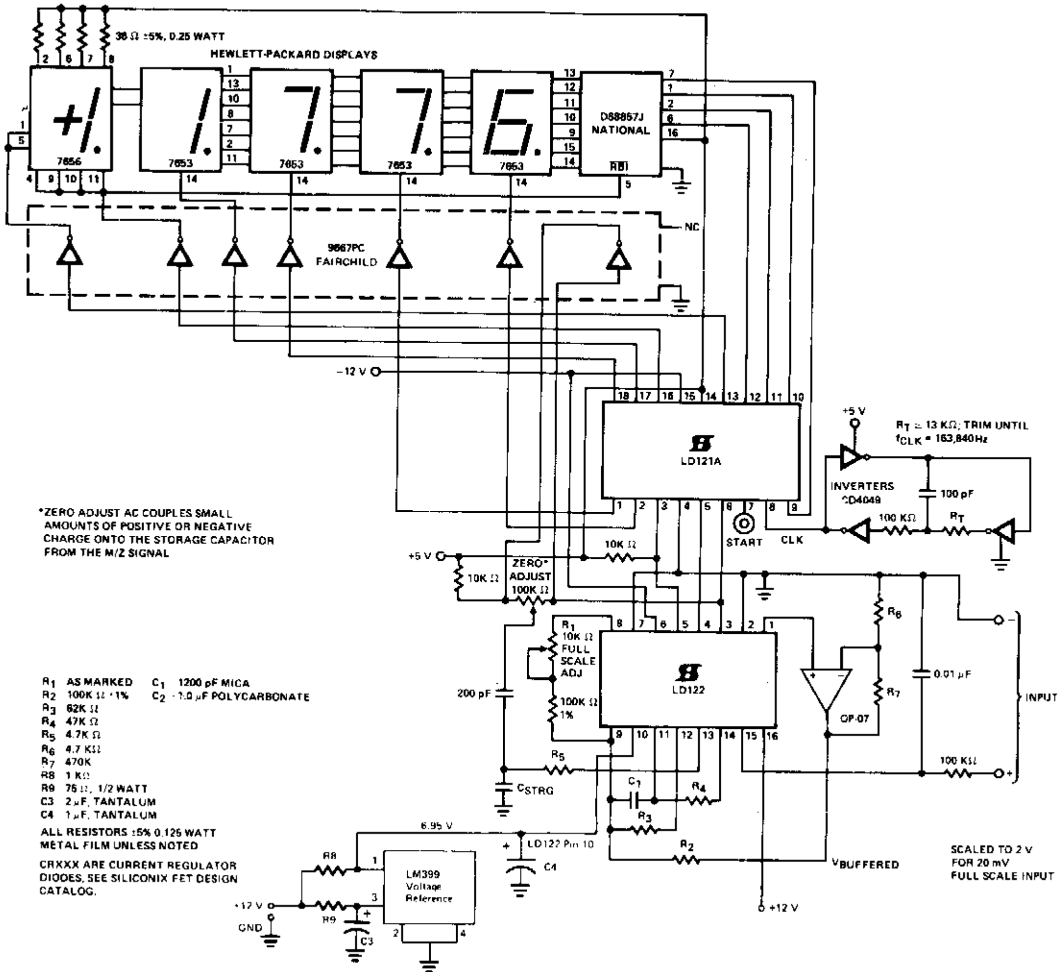
Output States	Condition V_{refA} V_{refB}
$V_{in} < V_{refB}$	Red
$V_{refB} < V_{in} < V_{refA}$	Orange/ Yellow
$V_{in} > V_{refA}$	Green

ELECTRONIC ENGINEERING

Fig. 124-1

A tricolor LED, acts as the visual indicator of the voltage level. The voltage to be measured is connected to the two comparators in parallel. The first 20-KΩ trimmer defines a voltage between ±5 V and this becomes the full-scale value of the reference voltage. The second trimmer is a fine adjustment to give any reference voltage between 0 V and the full-scale voltage. Thus, it is possible to select both positive and negative reference voltages. During the initialization procedure, a voltage, equal to the reference voltage of each comparator, is connected to the input terminal, and the offset balance potentiometer is adjusted to give a reading between the high and low output voltage levels. The inverter following comp A ensures that, whatever the input voltage, at least one diode is lit. The two inverters following comp B leave the voltage largely unchanged, but provide the current necessary to illuminate the diode. The value of the resistance should be chosen so that the current through any single diode does not exceed the specified limit, usually 30 mA. The LED contains a red and a green diode with a common cathode. When both diodes are lit, a third color, orange, is emitted. With $V_{\mathrm{ref\,A}}$ greater than $V_{\mathrm{ref\,B}}$, the output states given in the diagram apply.

4½-DIGIT DVM

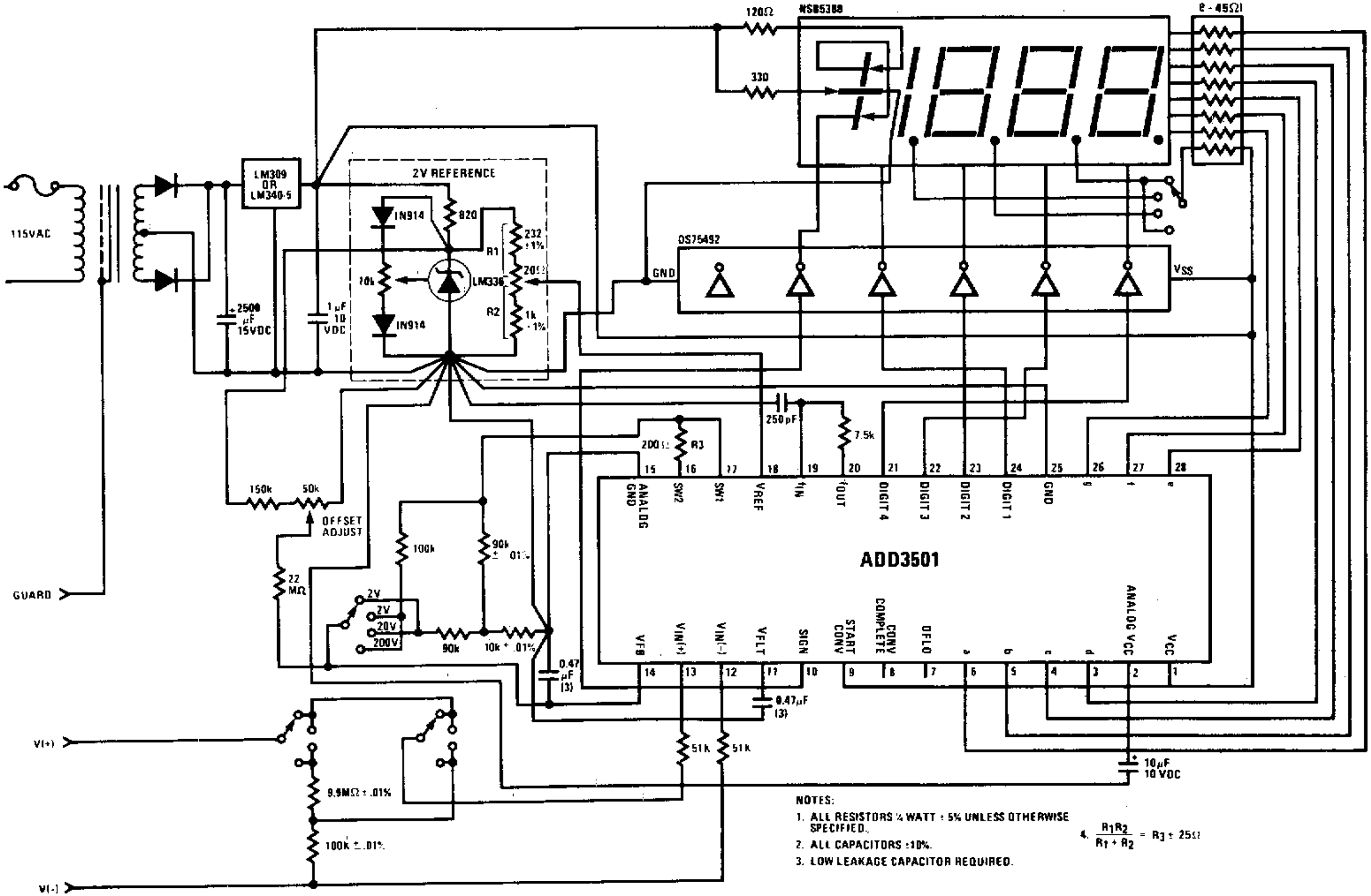


SILICONIX

Fig. 124-2

- 1-μV resolution
- Overrange blinking
- 0 – 19.999 mV input voltages
- Zero adjust-to-null offset introduced by PC board leakage and the comparator.

FULL-SCALE FOUR-DECADE 3½-DIGIT DVM

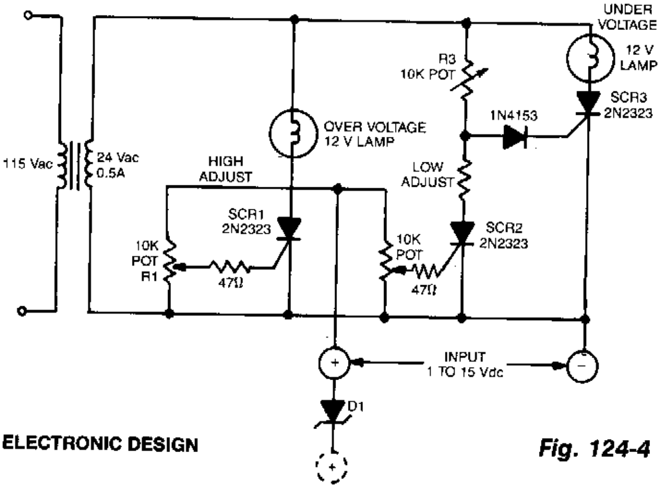


NATIONAL SEMICONDUCTOR CORP.

Fig. 124-3

This DVM circuit uses a National ADD3501 DVM chip and an LM336 reference IC to create a simple DVM with relatively few components. When making a single range panel meter, the range switching components can be left out, as required.

OVER/UNDER VOLTAGE MONITOR

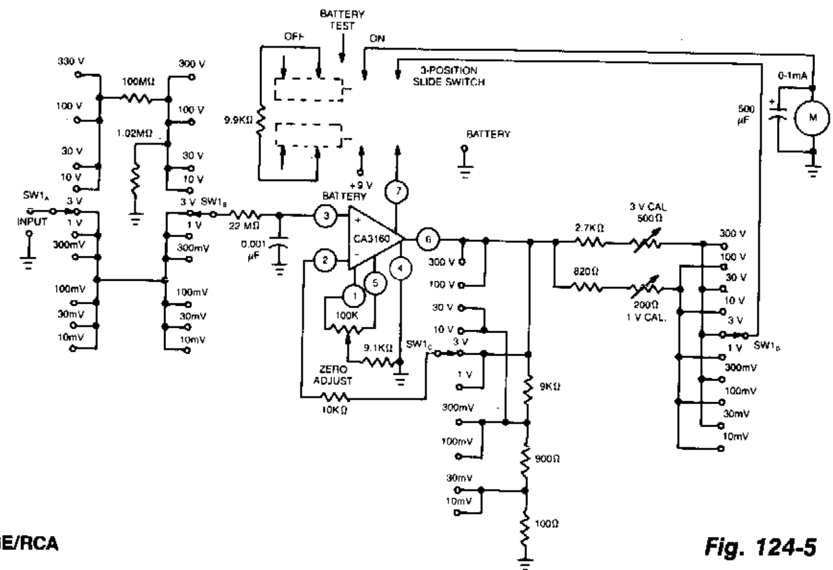


ELECTRONIC DESIGN

Fig. 124-4

Any potential from 1 to 15 V can be monitored with this circuit. Two lamps alert any undesirable variation. The voltage differential from lamp turn-on to turn-off is about 0.2 V at any setting. High and low set points are independent of each other. The SCRs used in the circuit should be the sensitive gate type. R3 must be experimentally determined for the particular series of SCRs used. This is done by adjusting R3 to the point where the undervoltage lamp turns on when no signal is present at the SCR2 gate. Any 15-V segment can be monitored by putting the zener diode, D1, in series with the positive input lead. The low set-point voltage will then be the zener voltage plus 0.8 V.

HIGH INPUT RESISTANCE DC VOLTMETER

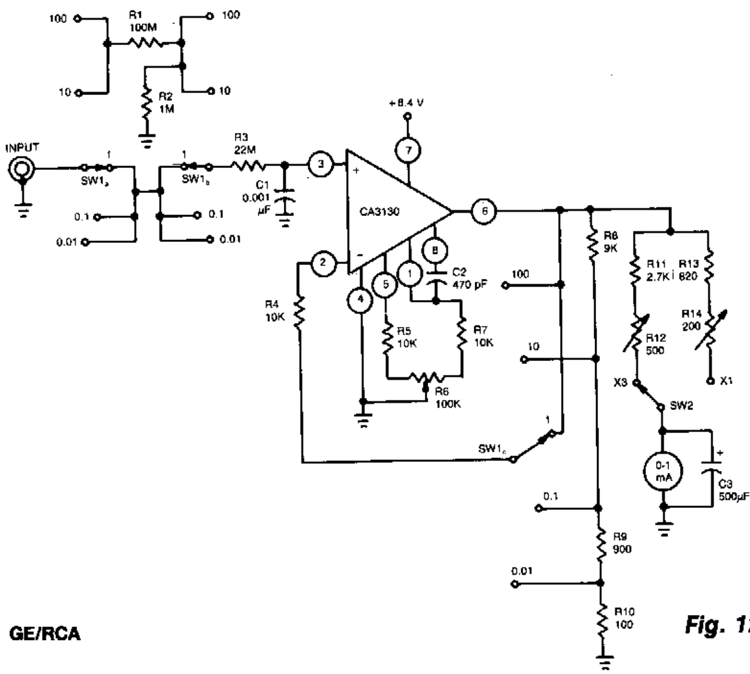


GE/RCA

Fig. 124-5

This voltmeter exploits a number of the CA3160 BiMOS op amp's useful characteristics. The available voltage ranges from 10 mV to 300 V. Powered by a single 8.4-V mercury battery, this circuit, with zero input, consumes approximately 500 μA. Thus, at full-scale input, the total supply current will increase by 1000 μA.

DC VOLTMETER



GE/RCA

Fig. 124-6

This dc voltmeter, with high input resistance, uses a CA3130 BiMOS op amp and measures voltages from 10 mV to 300 V. Resistors R12 and R14 are used individually to calibrate the meter for full-scale deflection. Potentiometer R6 is used to null the op amp and meter on the 10-mV range by shorting the input terminals, then adjusting R6 for the first indication of upscale meter deflection.

VOLTAGE FREEZER

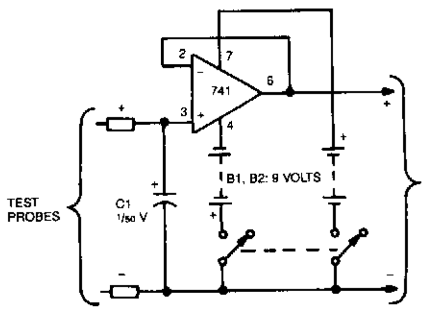


Fig. 124-7

This circuit reads and stores voltages, thus freezing the meter reading even after the probes are removed. The op amp is configured as a unity-gain voltage follower, with C1 situated at the input to store the voltage. For better performance, use an LF13741 or a TL081 op amp in place of the 741. These two are JFET devices and offer a much higher input impedance than the 741.

MULTIPLEXED COMMON-CATHODE LED-DISPLAY ADC

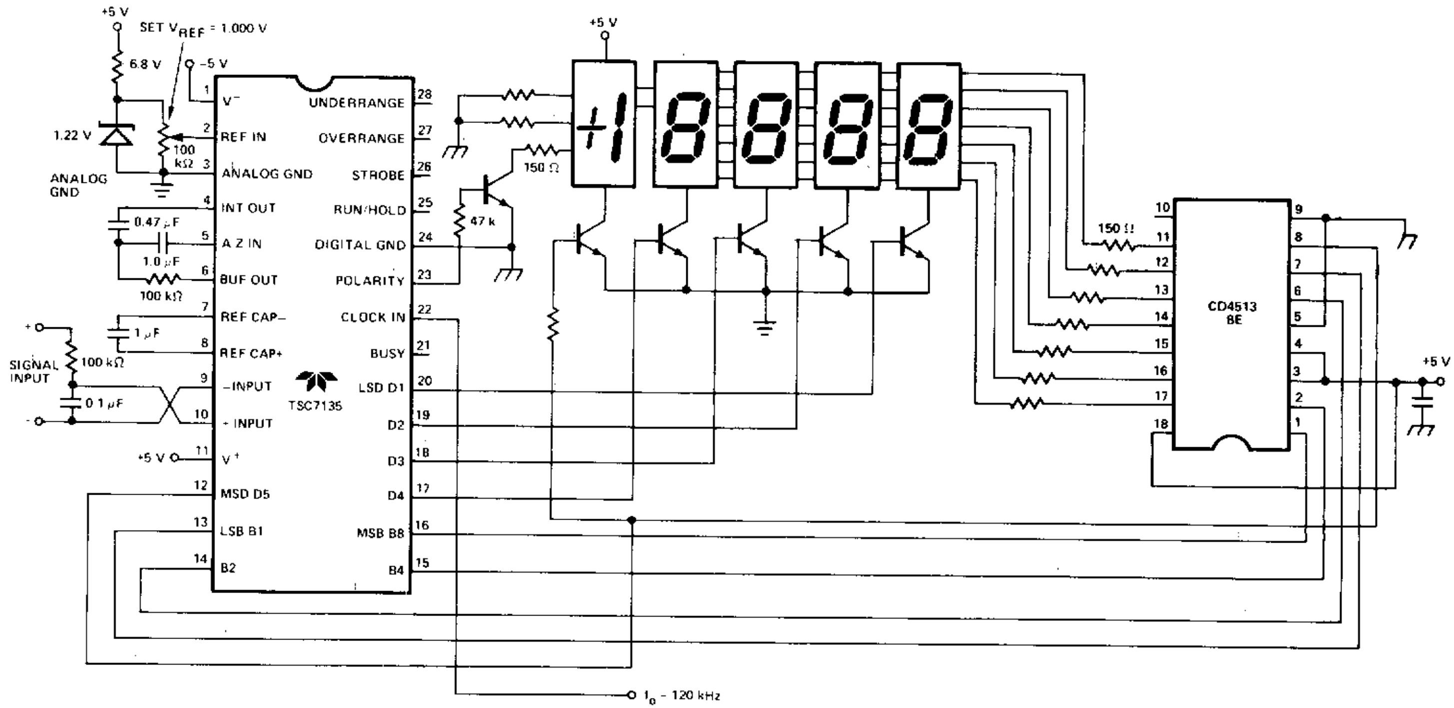


Fig. 124-8

Here, a Teledyne TSC7135 DVM chip is used to drive a multiplexed 5-digit display. A CD4513BE CMOS IC, for common cathode drive, is used as a segment driver selected by pins 17–20 of the DVM chip. The transistors can be any suitable npn type such as 2N3904, etc.

AC VOLTMETER

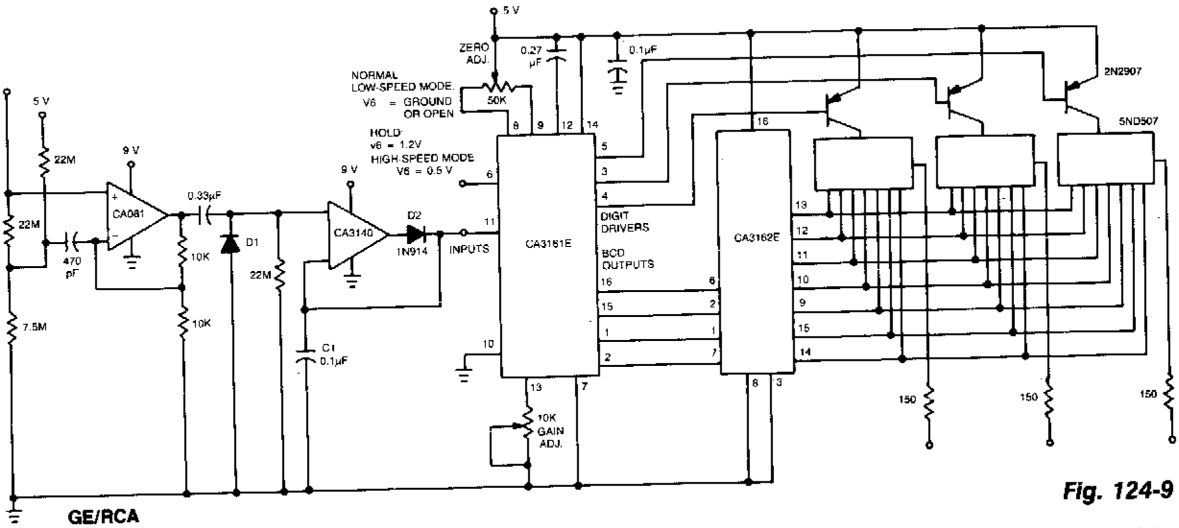


Fig. 124-9

GE/RCA

CA081 and CA3140 BiMOS op amp offer minimal loading on the circuits being measured. The wide bandwidth and high slew rate of the CA081 allow the meter to operate up to 0.5 MHz.

FET VOLTMETER

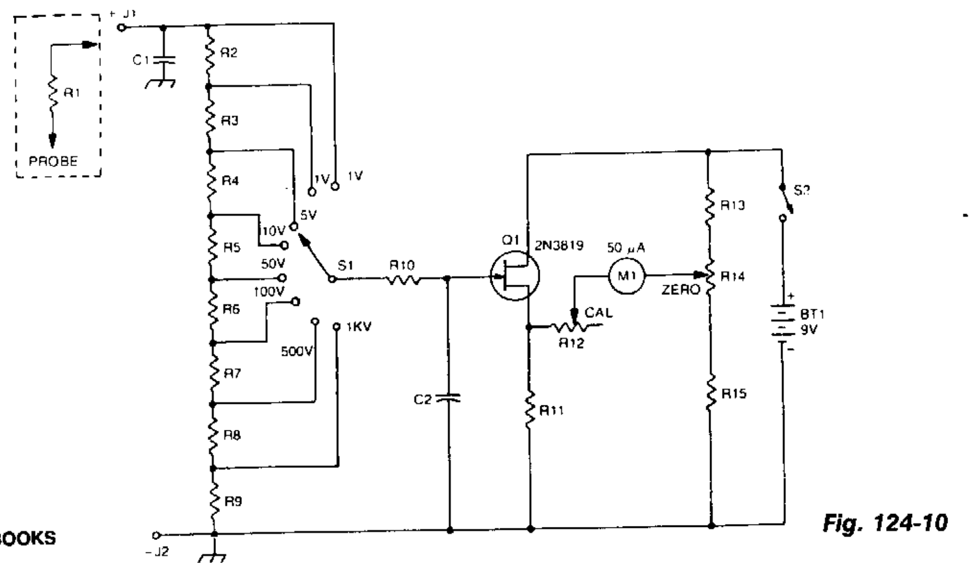
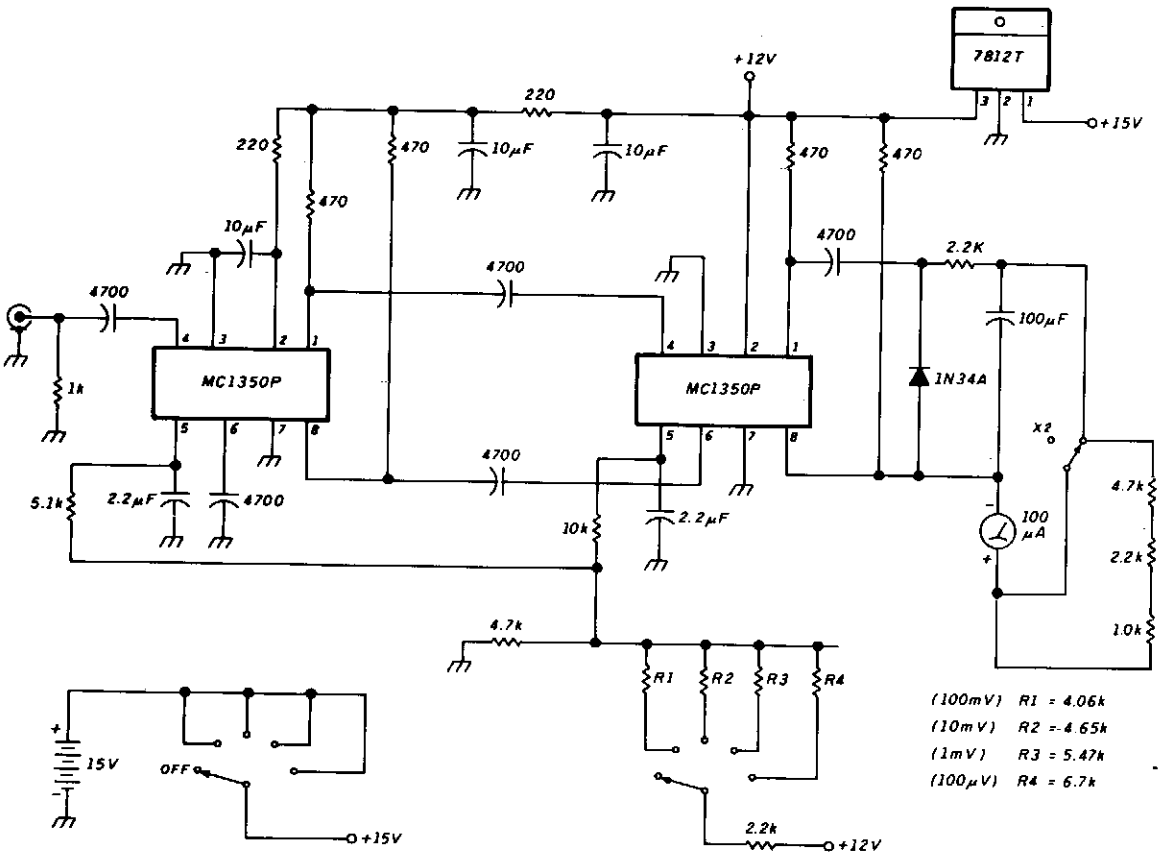


Fig. 124-10

TAB BOOKS

A 2N3819 FET provides a *solid-state VOM*. The 2N3819 acts as a *cathode follower* in a VOM. The bias offset (meter null) is obtained with R14 and R12 sets full-scale calibration. R2 through R9 should total about 10 MΩ. R10 is a protective resistor, and C2 provides ac bypassing to limit rf and noise pickup.

SENSITIVE RF VOLTMETER

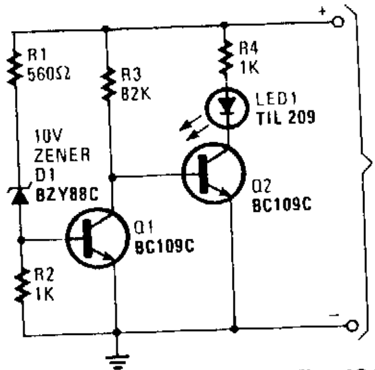


HAM RADIO

Fig. 124-11

This schematic shows a peak-reading diode voltmeter driven by two stages of amplification. A 100-μF capacitor provides a fairly large time constant, which results in satisfactory meter damping. The limited differential output voltage coupled with an overdamped meter prevents most *needle pinning* when you select an incorrect range position, or make other errors. An SPST toggle switch selects additional series resistance. This X2 function gives some more overlap of the sensitivity ranges. The resistance values shown are correct for use with a 100-μA meter with 1500-Ω internal resistance.

VOLTAGE MONITOR

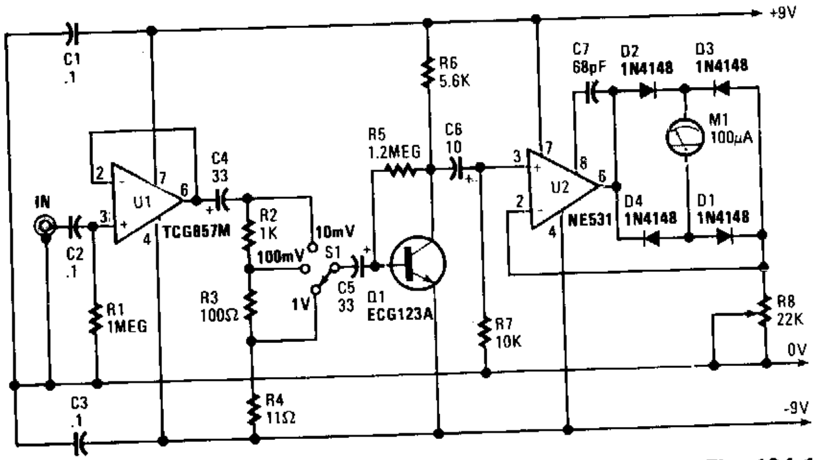


HANDS-ON ELECTRONICS

Fig. 124-12

If the battery voltage exceeds about 11 V, current flows through R1, D1, and R2. The voltage produced as a result of current flow through R2 is sufficient to bias transistor Q1 into conduction. That places the collector voltage of Q1 virtually at ground. Therefore, Q2, driven from the collector of Q1, is cut off, LED1 and current-limiting resistor R4 are connected in the collector circuit of Q2. With Q2 in the cut-off state, the LED does not light. Should Q1's base voltage drop below approximately 0.6 V, Q1 turns off, biasing Q2 on and illuminating LED1 to indicate that the battery voltage has fallen below the 11 V threshold level.

AUDIO MILLIVOLTMETER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

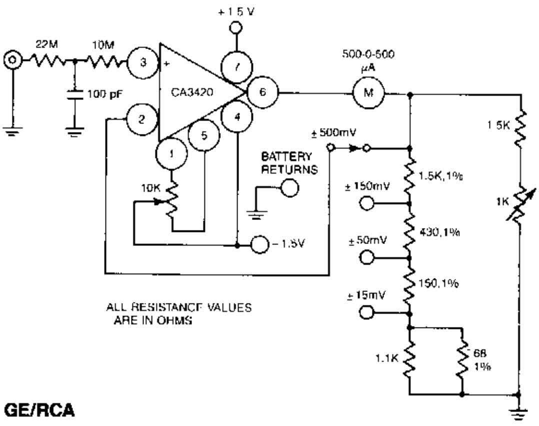
Fig. 124-13

Capacitor C4 couples the output of U1 to a simple attenuator, which is used to provide a loss of 0 dB, 20 dB, or 40 dB, depending on the setting of range switch S1. The circuits sensitivity is 10-V rms for full-scale deflection, so the attenuator gives additional ranges of 100-mV and 1-V rms. The attenuator output is connected through capacitor C5 to common-emitter amplifier Q1, which has a high-voltage gain of 40 dB.

To get linear scaling on the meter, we have to use an active-rectifier circuit built around U2. That IC is connected so that its noninverting input is biased to the 0-V bus via R7. Capacitor C6 couples the output of Q1 to the noninverting input of U2; C7 is the compensation capacitor for U2.

The voltage gain of U2 is set by the difference in resistance between the output and the inverting input, and between the inverting input and the ground bus. One resistance is made up of the diode-bridge rectifier D1 through D4, the other by resistor R8. This circuit has a nearly flat frequency response to about 200 kHz.

HIGH-INPUT RESISTANCE VOLTMETER



GE/RCA

Fig. 124-14

A resistance of 1,000,000 MΩ takes advantage of the high input impedance of the CA3420 BiMOS op amp. Only two 1.5-V AA-type penlite batteries are required for use. Full-scale deflection is ±500 mV, ±150 mV, and ±15 mV.

FREQUENCY COUNTER

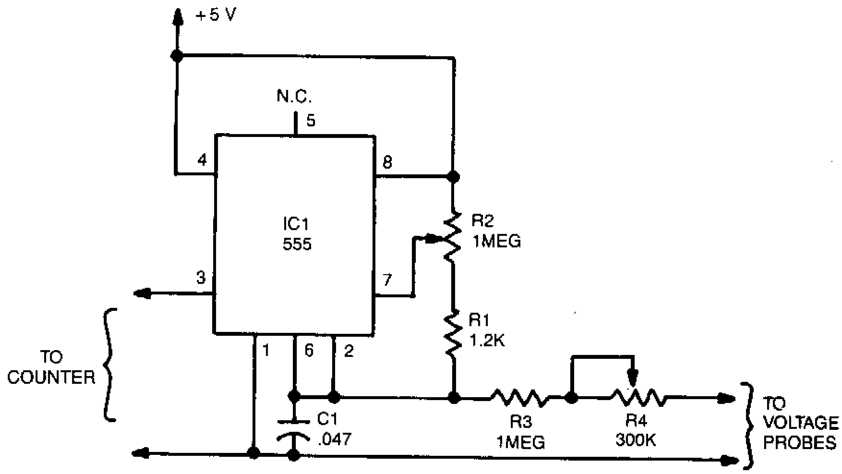
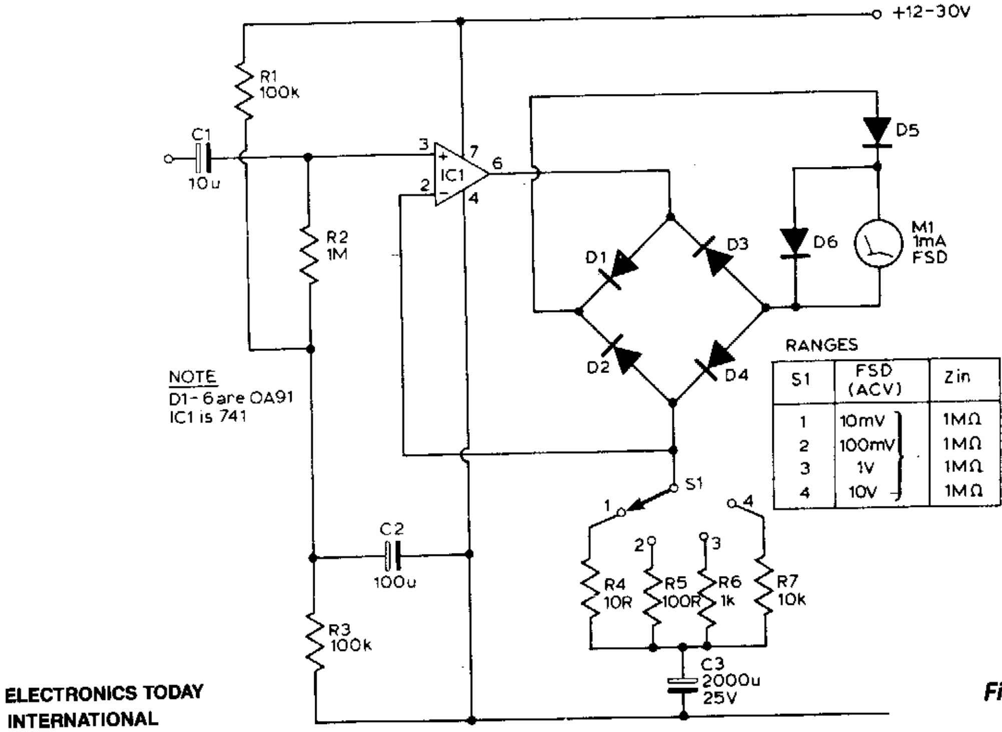


Fig. 124-15

The output frequency from IC pin 3 is determined by the voltage input to pin 6. A standard frequency counter can be used to measure voltages directly over a limited range from 0 to 5 V. In this circuit, the 555 is wired as an astable multivibrator. Resistor R2 determines the output frequency when the input to the circuit (the voltage measured by the voltage probes) is zero. R4 is a scaling resistor that adjusts the output frequency so that a change in the input voltage of 1 V will result in a change in the output frequency of 10 Hz. That will happen when the combined resistance of R3 and R4 is 1.2 MΩ. To calibrate short the voltage probes together, adjust R2 until the reading on the frequency counter changes to 00 Hz. Then, use the voltage probes to measure an accurate 5-V source and adjust R4 until the frequency counter reads 50 Hz.

AUDIO MILLIVOLTMETER



S1	FSD (ACV)	Zin
1	10mV	1MΩ
2	100mV	1MΩ
3	1V	1MΩ
4	10V	1MΩ

ELECTRONICS TODAY INTERNATIONAL

Fig. 124-16

This circuit has a flat response from 8 Hz to 50 kHz at −3 db on the 10-mV range. The upper limit remains the same on the less sensitive ranges, but the lower frequency limit covers under 1 Hz.

LOW-VOLTAGE INDICATOR

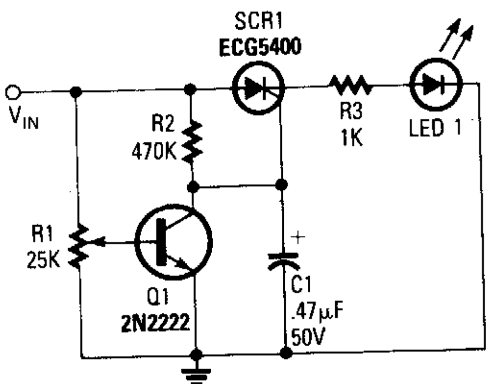
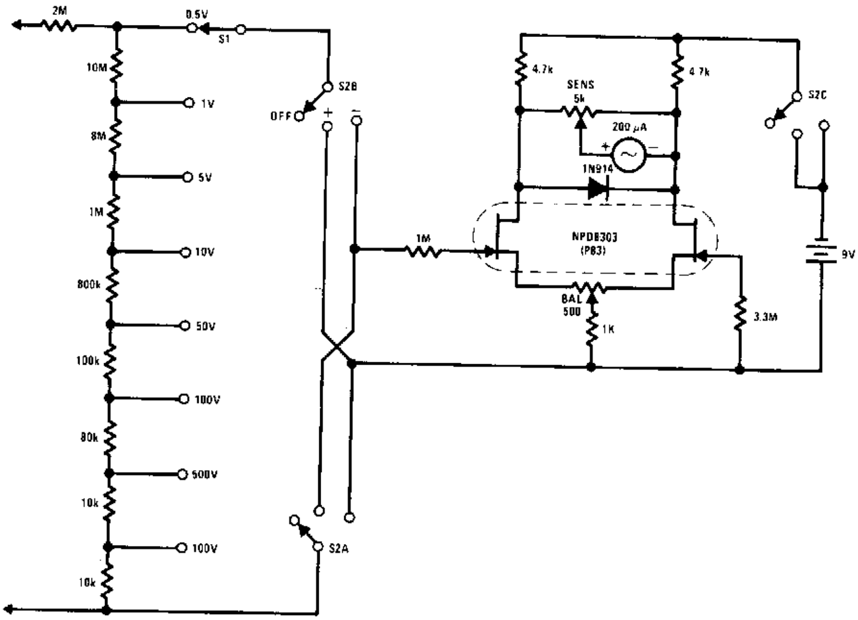


Fig. 124-17

Input terminal V_{IN} is connected to the +V line of the circuit that the indicator is to monitor, and the grounds of both circuits are connected together. The position of potentiometer R1's wiper determines Q1's base voltage. As long as the transistor gets enough bias voltage to remain on, the low voltage at the collector will keep the SCR from firing. As the battery voltage starts to fall, the transistor's base voltage will fall as well. When Q1 turns off (V_{IN} drops), the collector voltage increases. That voltage provides enough gate drive to turn on the SCR, which turns on the LED. The LED could also be a buzzer or almost any other type of warning device.

FET VOLTMETER

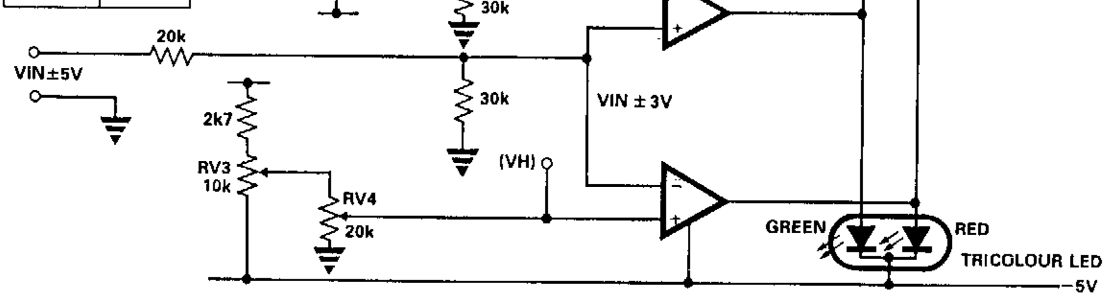


NATIONAL SEMICONDUCTOR CORP. *Fig. 124-18*

This FETVM replaces the function of the VTVM and rids the instrument of the usual line cord. In addition, FET drift rates are far superior to vacuum tube circuits, allowing a 0.5 V full-scale range which is impractical with most vacuum tubes. The low leakage, low noise NPD8303 is ideal for this application.

SIMPLIFIED VOLTAGE-LEVEL SENSOR

INPUT	OUTPUT
VIN*>VH	GREEN
VL<VIN*<VH	YELLOW
VIN*<VL	RED



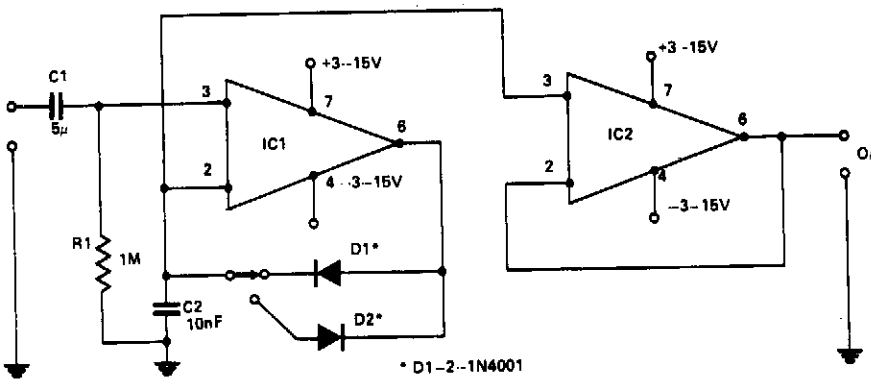
ELECTRONIC ENGINEERING *Fig. 124-19*

This circuit uses only one IC, either 1, LM393 dual comparator or 1/2, LM339 quad comparator. RV1 and RV3 set the full scale reference voltage, and RV2 and RV4 set the switching thresholds to a value between 0 V and the full-scale reference. The change in input voltage needed to fully switch the output state is less than 0.05 mV (typical).

An alternative is:

INPUT	OUTPUT
V_{IN}, V_H	red
V_H, V_{IN}, V_L	yellow
V_{IN} V_L	green

PEAK PROGRAM DETECTOR

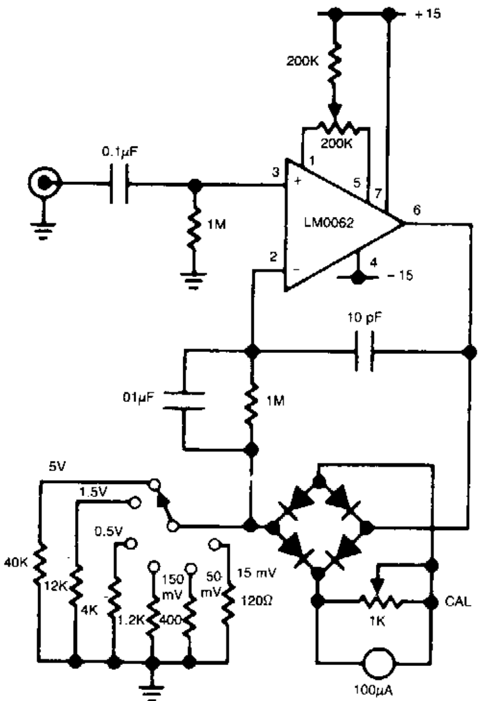


ELECTRONICS TODAY INTERNATIONAL

Fig. 124-20

This circuit will allow a multimeter to display the positive or negative peaks of an incoming signal. A 741, IC1, is used in the noninverting mode with R1 defining the input impedance. D1 or D2 will conduct on a positive or negative peak, charging C2 until the inverting input is at the same dc level as the incoming peak. This level will maintain the voltage until a higher peak is detected, then this will be stored by C2. Another 741, IC2, prevents loading by the multimeter. Connected in the noninverting mode as a unity gain buffer, output impedance is less than 1 Ω. This circuit has a useful frequency response from 10 Hz to 100 kHz at ±1 dB. High linearity is ensured by placing the diodes in the feedback loop of IC1, effectively compensating for the 0.6 V bias that these components require.

WIDE-RANGE AC VOLTMETER

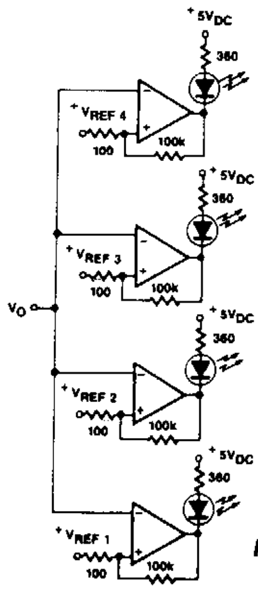


NATIONAL SEMICONDUCTOR CORP.

In this circuit, a diode bridge is used as a meter rectifier. The offset voltage is compensated for by the op amp, since the bridge is in the feedback network.

Fig. 124-21

VISABLE VOLTAGE INDICATOR



SIGNETICS

Fig. 124-22

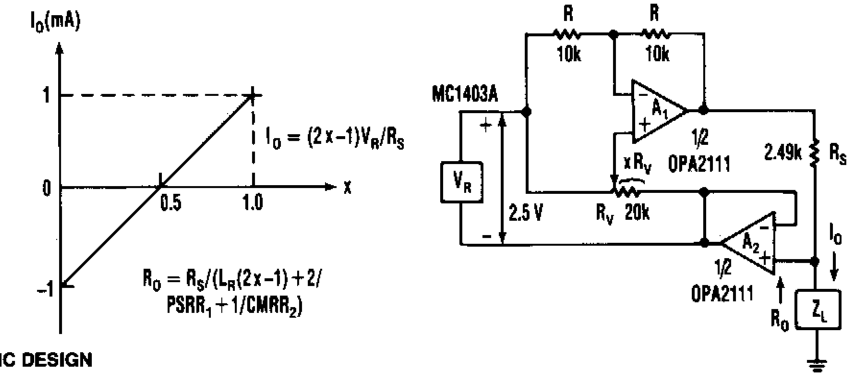
125

Voltage References

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

Bipolar Reference Source
Expanded-Scale Analog Meter
Digitally Controlled Voltage Reference

BIPOLAR REFERENCE SOURCE



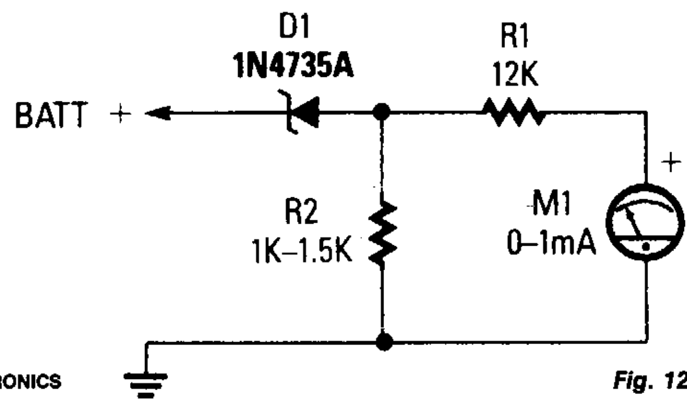
ELECTRONIC DESIGN

Fig. 125-1

This current source has continuous control of the magnitude and polarity of its amplifier gain and needs only one voltage reference. The circuit includes reference V_R, voltage-amplifier circuit A1 with gain-setting resistor R_S, and bootstrap-follower amplifier A2. The bootstrapping converts the circuit to a current source and allows the load to be grounded. Any voltage developed across load Z_L feeds back to the reference and voltage amplifier, making their functions immune to that voltage. Then the current-source circuitry floats, instead of the load.

The voltage reference is connected to both the inverting and noninverting inputs of A1; this provides a balanced combination of positive and negative gain. The inverting connection has equal feedback resistors, R, for a gain of -1, and the noninverting connection varies according to the fractional setting, X, of potentiometer R_V. X controls the noninverting gain and adjusting it counters the effect of some of the inverting gain. The value of X is the portion of R_V's resistance from the noninverting input of A1 to the temporarily grounded output of A2. Between potentiometer extremes, the current varies with X ± 1 mA.

EXPANDED-SCALE ANALOG METER

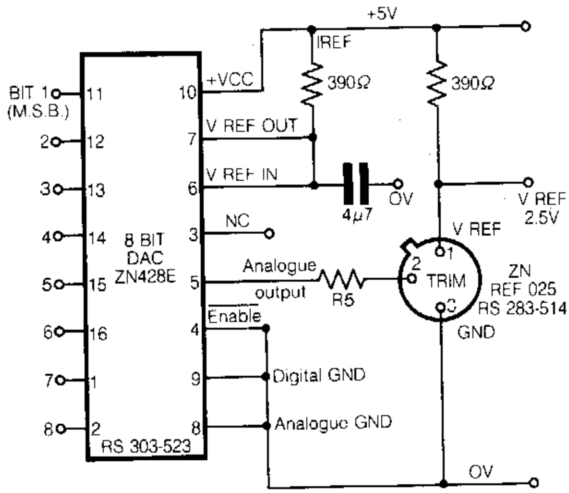


POPULAR ELECTRONICS

Fig. 125-2

The circuit consists of 0 – 1 mA meter M1, 6.2-V zener diode D1, and 12-KΩ, 1% resistor R1. R2 is included in the circuit as a load resistor for the zener diode. The value of R2 isn't critical; use a value of 1000 to 1500 Ω. The meter reads from 6 to 18 volts, which is perfect for checking a car's charging system.

DIGITALLY-CONTROLLED VOLTAGE REFERENCE



ELECTRONIC ENGINEERING

Fig. 125-3

This circuit shows a simple method of achieving a voltage reference which can be trimmed using an 8-bit DAC with an on-chip voltage reference. The analogue output from the DAC drives the trim pin of the ZNREF025 via a resistor, R_S. When $R_S = 0$, this circuit will produce a trim range exceeding $\pm 5\%$ of the nominal reference voltage. When R_S is greater than zero, the trim range is reduced. It was found that after dividing the trim range by two, the needed value for R_S was approximately 1 MΩ.

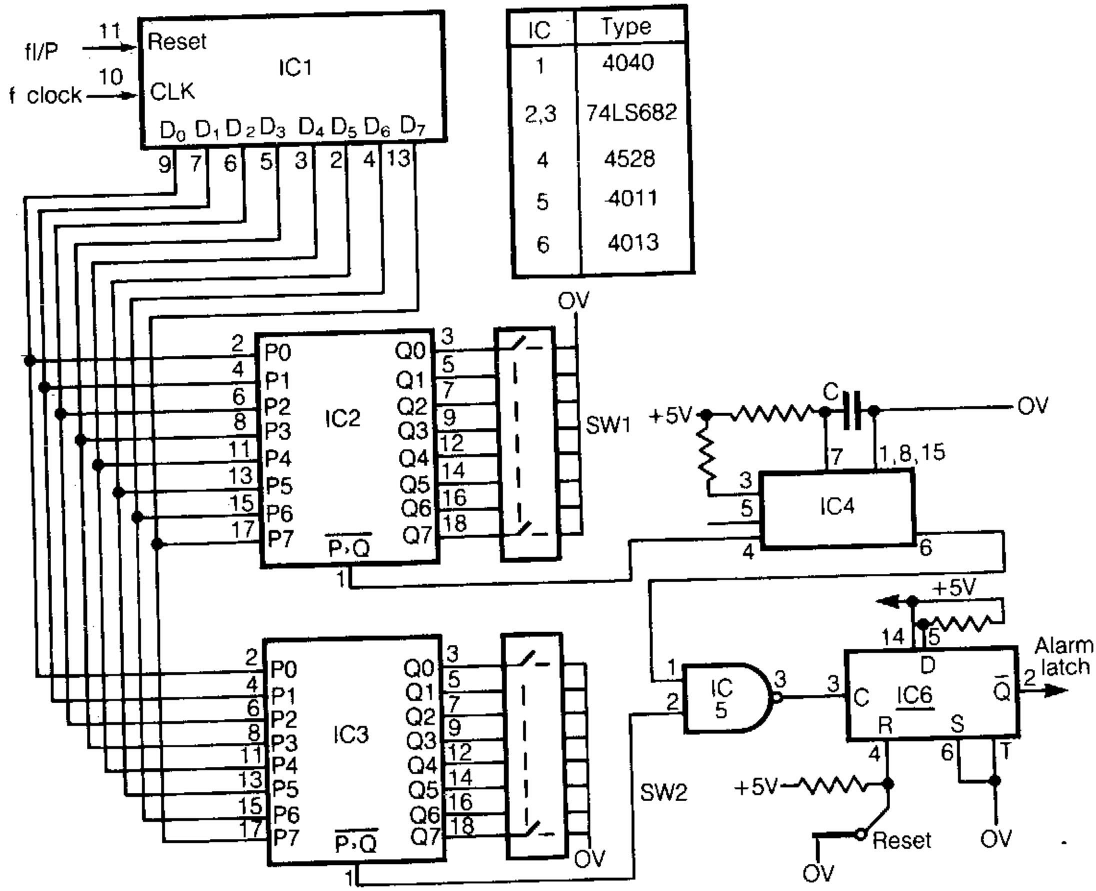
The reference voltage of the ZNREF025 can be set to an accuracy determined by the trim range of the device itself and the accuracy of the DAC. Increasing R_S reduces the percentage trim range and hence increases the effective voltage resolution. Other voltage references from the ZNREF series can be used with this circuit if other voltages are required. This voltage reference can be used to set the value of V_{REF} to a much tighter tolerance, than the data sheet specification of $\pm 1\%$, in a much wider range of operating conditions. Applications could include any system with automatic self-calibration of instrumentation, such as in electronic weighing scales.

126

Window Detectors/ Comparators/Discriminators

The sources of the following circuits are contained in the Sources section beginning on page 782. The figure number contained in the box of each circuit correlates to the sources entry in the Sources section.

DIGITAL FREQUENCY WINDOW



IC	Type
1	4040
2,3	74LS682
4	4528
5	4011
6	4013

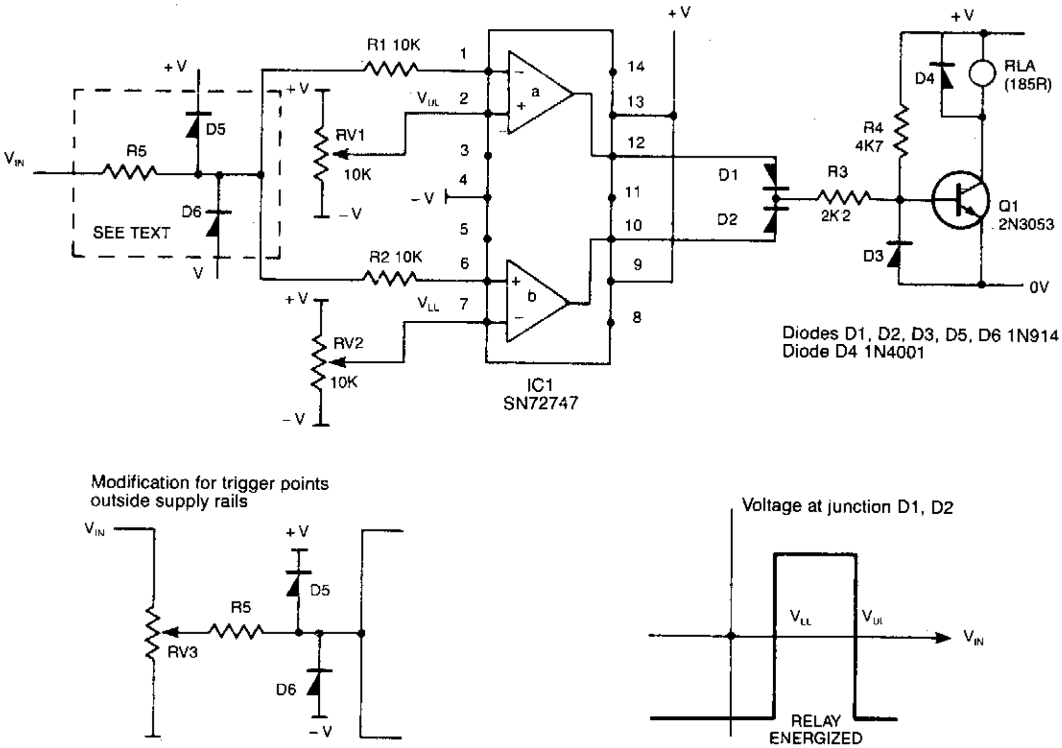
ELECTRONIC ENGINEERING

Fig. 126-1

This circuit detects frequency variation above or below preset limits. IC1 is a binary counter clocked at F_{CLK}. The outputs are compared with switch preset values by IC2 and IC3. The input signal, which must be a positive-going pulse, is used to reset IC1. The *P greater than Q* output of the comparators is at logic *O* for input frequencies below the preset values. Above the preset count, a pulse train is output.

IC2, detects a low input by supplying the pulse train to a retriggerable monostable, IC4. When the input frequency falls below the preset value in SW1, the monostable is no longer triggered and its output falls to logic *O*. IC3 detects the frequency high state SW2, and outputs directly when this occurs. The outputs from both comparators can then be latched as shown, using IC5 and IC6. The clock frequency is related to input and switch values: switch value = $F_{CLK/input}$. The time constant of IC4 is not critical, but must obviously exceed the maximum input pulse period.

WINDOW DETECTOR

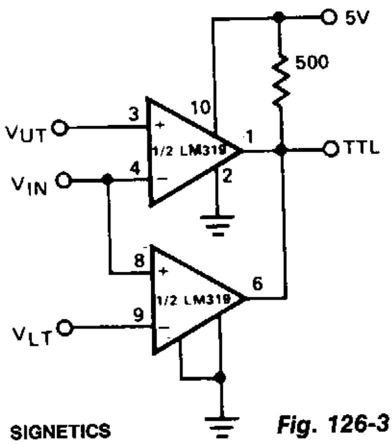


ELECTRONICS TODAY INTERNATIONAL

Fig. 126-2

This circuit de-energizes a normally energized relay if the input voltage goes above or below two individually set voltages. The transistor driving the relay is normally turned on by R4, so the relay is normally energized. If the cathode of D1 or D2 is taken negative, Q1 will turn off and the relay will de-energize. The IC is a 72747 dual op amp used without feedback, so the full gain of about 100dB is available. The amplifier output will thus swing from full positive to full negative for a few mV change at the input. The relay is therefore only energized if V_{IN} is between V_{UL} and V_{LL}. The two limits can be set anywhere between the supply rails, but obviously V_{UL} must be more positive than V_{LL}. If V_{IN} can go outside the supply rails, D5, D6, and R5 should be added to prevent damage to IC1. If V_{UL} and V_{LL} are required to be outside the supply rails, V_{IN} can be reduced by RV3. The supplies can be any value, providing that the voltage across them is not more than 30 V.

WINDOW DETECTOR

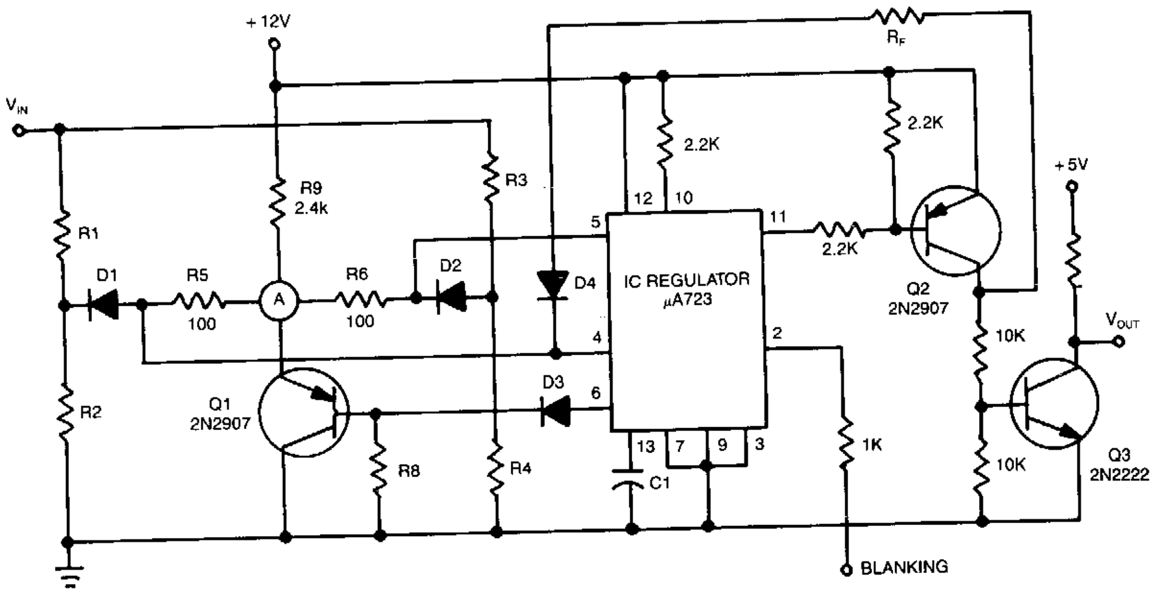


$V_{OUT} = 5V$ for $V_{LT} < V_{IN} < V_{UT}$
$V_{OUT} = 0$ for $V_{IN} < V_{LT}$ or $V_{IN} > V_{UT}$

SIGNETICS

Fig. 126-3

WINDOW DETECTOR

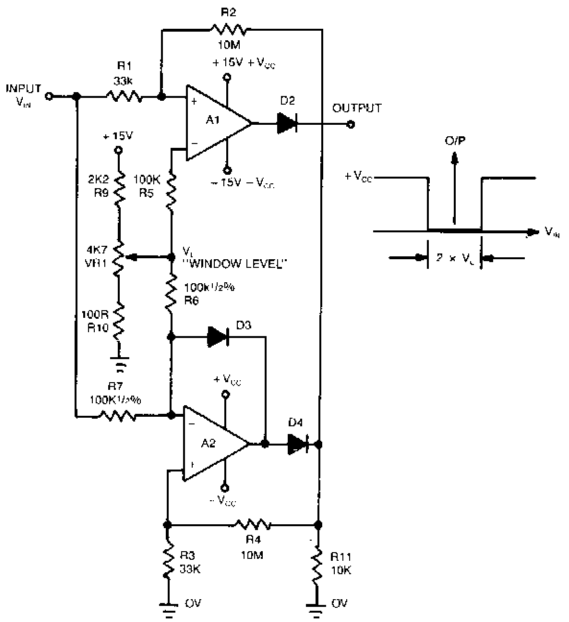


EDN

Fig. 126-4

The detector circuit compares the output voltage of two separate voltage dividers with a fixed reference voltage. The resultant absolute error signal is amplified and converted to a logic signal that is TTL compatible.

WINDOW DETECTOR

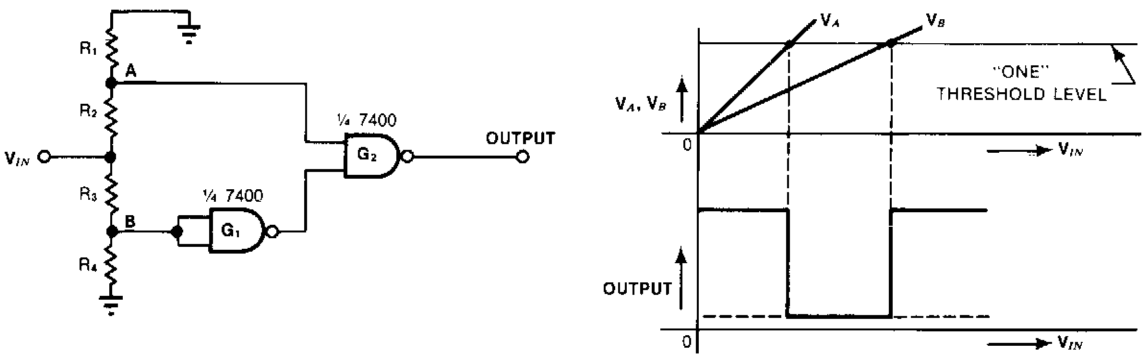


This novel window detector uses only two op amps. The width of the window can be changed by the 4.7-KΩ potentiometer.

ELECTRONIC ENGINEERING

Fig. 126-5

SIMPLE WINDOW DETECTOR

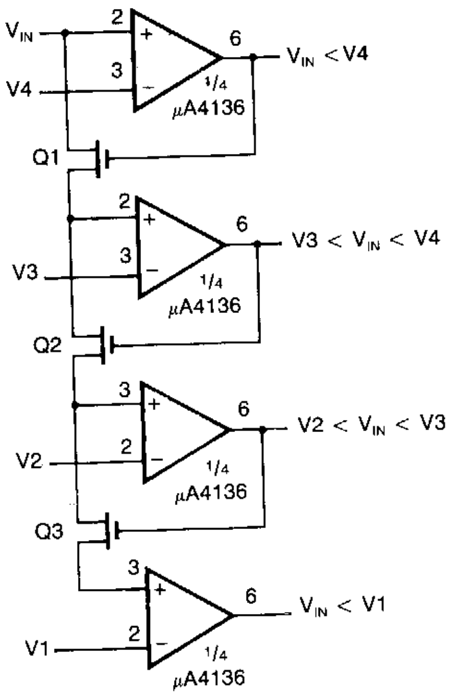


EDN

Fig. 126-6

This simple window detector uses only half of a 7400 quad NAND gate plus four resistors, chosen so that the voltage at point A exceeds the voltage at point B for any input voltage. With no input applied or when V_{IN} is at ground, the output of gate G1 is one; hence G2's output is also one. As the input voltage increases, V_A rises faster than V_B. When V_A reaches an acceptable one level, the circuit's output drops to zero. As the input continues to increase, V_B rises to an acceptable level, changing the output of G2 to one.

MULTIPLE APERTURE WINDOW DISCRIMINATOR



The circuit shown here uses μA4136 comparators and FETs Q1 through Q3.

FAIRCHILD CAMERA AND INSTRUMENT CORP.

Fig. 126-7

Sources Index

Chapter 1

Fig. 1-1. Radio-Electronics, 2/89, p.51.
Fig. 1-2. QST, 2/86, p.22.

Chapter 2

Fig. 2-1. Popular Electronics, 8/89, p. 24.
Fig. 2-2. Reprinted from EDN, 5/16/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 2-3. Reprinted with permission from Electronic Design. Copyright 1988, Penton Publishing.
Fig. 2-4. Reprinted from EDN, 9/20/84, Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 2-5. Popular Electronics, 8/89, p. 22.
Fig. 2-6. Reprinted from EDN, 12/74, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 3

Fig. 3-1. Harris, Analog Product Data Book, 1988, p.10-167.
Fig. 3-2. Reprinted from EDN, 1/79, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 3-3. Electronic Engineering, Applied Ideas, 9/88, p. 25.
Fig. 3-4. Maxim, Maxim Advantage, p. 44.
Fig. 3-5. Harris, Analog Product Data Book, 1988, p. 10-95.
Fig. 3-6. Harris, Analog Product Data Book, 1988, p. 10-96.
Fig. 3-7. Hands-on Electronics, Fact Card No. 29.
Fig. 3-8. Hands-on Electronics, Fact Card No. 29.
Fig. 3-9. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 13.
Fig. 3-10. Signetics, 1987 Linear Data Manual Vol 2: Industrial, 11/6/86, p. 4-136.
Fig. 3-11. Signetics, 1987 Linear Data Manual Vol 2: Industrial, 2/87, p. 4-243.
Fig. 3-12. Siliconix, Integrated Circuits Data Book, 1988, p. 13-166.
Fig. 3-13. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 13.

Fig. 3-14. Harris, Analog Product Data Book, 1988, p. 10-110.
Fig. 3-15. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN18-3.
Fig. 3-16. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN21-5.
Fig. 3-17. Harris, Analog Product Data Book, 1988, p. 10-150.
Fig. 3-18. Reprinted from EDN, 9/1/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 3-19. Intersil, Applications Handbook, 1988, p. 3-181.
Fig. 3-20. Maxim, Maxim Advantage, p. 44.
Fig. 3-21. Electronic Engineering, 9/88/ p. 28.

Chapter 4

Fig. 4-1. Siliconix, Integrated Circuits Data Book, 1988, p. 6-148.
Fig. 4-2. Harris, Analog Product Data Book, 1988, p. 10-48.
Fig. 4-3. Signetics, 1987 Linear Data Manual Vol 2: Industrial, 2/87, p. 5-311.
Fig. 4-4. Maxim, Seminar Applications Book, 1988/89, p. 38.

Chapter 5

Fig. 5-1. Reprinted from EDN, 4/77, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 6

Fig. 6-1. Harris, Analog Product Data Book, 1988, p. 10-173.
Fig. 6-2. Harris, Analog Product Data Book, 1988, p. 10-13.
Fig. 6-3. Signetics, RF Communications Handbook, 1989, p. 2-14 and 2-15.

Chapter 7

Fig. 7-1. Harris, Analog Product Data Book, 1988, p. 10-108.
Fig. 7-2. Signetics, Analog Data Manual, 1983, p. 10-20.
Fig. 7-3. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-153.
Fig. 7-4. Harris, Analog Product Data Book, 1988, p. 10-161.
Fig. 7-5. Hands-On Electronics, Summer 1984, p. 74
Fig. 7-6. Radio-Electronics, 8/88, p. 33.
Fig. 7-7. Signetics, RF Communications Handbook, 1989, p. 1-61 and 1-62.
Fig. 7-8. Popular Electronics, 7/89, p 26.
Fig. 7-9. Popular Electronics, Fact Card No. 110.
Fig. 7-10. Harris, Analog Product Data Book, 1988, p. 10-174.
Fig. 7-11. Harris, Analog Product Data Book, 1988, p. 10-174.
Fig. 7-12. Hands-On Electronics, 5/87, p. 96.
Fig. 7-13. QST, 1/89, p. 20.
Fig. 7-14. Hands-On Electronics/Popular Electronics, 11/88, p. 39.

Chapter 8

Fig. 8-1. Gernsback Publications Inc., 42 New Ideas, 1984, p. 9.
Fig. 8-2. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-158.
Fig. 8-3. Popular Electronics, 8/89, p. 22.
Fig. 8-4. Gernsback Publications Inc. 42 New Ideas, 1985, p. 28.
Fig. 8-5. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.
Fig. 8-6. ZeTeX, (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-155.
Fig. 8-7. Radio-Electronics, 7/85, p. 55.
Fig. 8-8. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p 3.
Fig. 8-9. Reprinted from EDN, 8/81, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 8-10. Intersil, Component Data Catalog, 1987, p. 2-108.
Fig. 8-11. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-67.
Fig. 8-12. Hands-On Electronics, 9–10/86, p. 27.
Fig. 8-13. TAB Books, The Giant Book of Easy-to-Build Electronic Projects, 1982, p. 186.
Fig. 8-14. TAB Books, 101 Sound, Light, and Power IC Projects, 1986, p. 139.
Fig. 8-15. Hands-On Electronics, 5/87, p. 95.
Fig. 16. Hands-On Electronics/Popular Electronics, 12/88, p. 25.
Fig. 8-17. Popular Electronics/Hands-On Electronics, 3/89, p. 36.
Fig. 8-18. Radio-Electronics, 4/87, p. 67.

Chapter 9

Fig. 9-1. Motorola, Motorola Thyristor Device Data, Series A 1985, p. 1-6-54.
Fig. 9-2. Reprinted from EDN, 7/21/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 9-3. Linear Technology Corp., Linear Databook, 1986, p. 4-15.
Fig. 9-4. Radio-Electronics, 7/86, p. 67.
Fig. 9-5. Motorola, Motorola Thyristor Device Data, Series A 1985, p. 1-6-46.
Fig. 9-6. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-164.
Fig. 9-7. Popular Electronics, 7/89, p. 81.
Fig. 9-8. Gernsback Publications, Inc., 42 New Ideas, 1984, p. 9.

Chapter 10

Fig. 10-1. Reprinted from EDN, 7/21/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 10-2. Popular Electronics/Hands-On Electronics, 5/89, p. 25.
Fig. 10-3. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 5.
Fig. 10-4. Maxim, 1986 Power Supply Circuits, p. 26.
Fig. 10-5. Hands-On Electronics, Spring 1985, p. 49.
Fig. 10-6. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S2-7.
Fig. 10-7. Reprinted from EDN, 9/5/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 10-8. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 6.
Fig. 10-9. Maxim, 1986 Power Supply Circuits, p. 121.
Fig. 10-10. Reprinted from EDN, 1/8/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 10-11. Linear Technology, Application Note 8, p. 2.
Fig. 10-12. GE, Optoelectronics, Third Edition, Ch. 6, p. 148.

Chapter 11

Fig. 11-1. Harris, Analog Product Data Book, 1988, p. 10-183.
Fig. 11-2. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.
Fig. 11-3. Harris, Analog Product Data Book, 1988, p. 10-106.
Fig. 11-4. Signetics, 1987 Linear Data Manual Vol 2: Industrial, 2/87, p. 4-78.
Fig. 11-5. Linear Technology, 1986 Linear Databook, p. 2-45.

Chapter 12

Fig. 12-1. Reprinted from EDN, 7/10/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 12-2. Ham Radio, 7/86, p. 88.

Chapter 13

Fig. 13-1. Intersil, Applications Handbook, 1988, p. 3-138.
Fig. 13-2. Hands-On Electronics, Fall 1984, p. 68

Chapter 14

Fig. 14-1. Radio-Electronics, 1/89, p. 55.
Fig. 14-2. Radio-Electronics, 2/89, p. 55
Fig. 14-3. Radio-Electronics, 2/89, p. 55
Fig. 14-4. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 5-36.

Chapter 15

Fig. 15-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 12/2/86, p. 6-78.
Fig. 15-2. Harris, Analog Product Data Book, 1988, p. 2-106.
Fig. 15-3. TAB Books, 44 Electronics Projects for the Darkroom, p. 101.
Fig. 15-4. Reprinted from EDN, 10/78, © 1989 Cahners Publishing Co,. a division of Reed Publishing USA.

Chapter 16

Fig. 16-1. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 3-46 and 3-47.
Fig. 16-2. Linear Technology, 1986 Liner Databook, p. 2-44.
Fig. 16-3. Reprinted from EDN, 4/75, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 16-4. Reprinted from EDN, 4/79, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 16-5. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S1-42.
Fig. 16-6. Linear Technology, 1986 Linear Databook, p. 2-29.
Fig. 16-7. Reprinted from EDN, 2/18/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 16-8. Harris, Analog Product Data Book, 1988, p. 10-16.

Chapter 17

Fig. 17-1. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 4-373 and 4-375.
Fig. 17-2. Signetics, RF Communications Handbook, 1989, p. 2-17 and 2-18.

Fig. 17-3. Signetics, RF Communications Handbook, 1989, p. 2-15 and 2-16.

Fig. 17-4. Signetics, RF Communications Handbook, 1989, p. 2-17 and 2-18.

Chapter 18

Fig. 18-1. Electronic Engineering, Applied Ideas, 4/86, p. 33.

Fig. 18-2. Siliconix, Integrated Circuits Data Book, 1988, p. 6-40 and 6-41.

Fig. 18-3. Hands-On Electronics/Popular Electronics, 11/88, p. 43.

Fig. 18-4. Linear Technology Corp., Linear Databook, 1986, p. 2-98.

Fig. 18-5. Siliconix, Integrated Circuits Data Book, 1988, p. 13-200.

Fig. 18-6. Siliconix, Integrated Circuits Data Book, 1988. p. 5-140.

Fig. 18-7. Reprinted from EDN, 4/28/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 18-8. Harris, Analog Product Data Book, 1988, 2-106.

Fig. 18-9. Reprinted from EDN, 10/27/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 18-10. Reprinted from EDN, 9/1/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 18-11. Reprinted from EDN, 6/22/89, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 18-12. Reprinted with permission from Electronic Design. Copyright 1988, Penton Publishing.

Fig. 18-13. Reprinted from EDN, 9/1/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 18-14. Intersil, Component Data Catalog, 1987, p. 13-22.

Fig. 18-15. Electronic Engineering, Applied Ideas, 10/88, p. 37.

Fig. 18-16. Electronic Engineering, Applied Ideas, 12/88, p. 22.

Chapter 19

Fig. 19-1. Teledyne Semiconductor, Data Acquisition IC Handbook, 1985, p. 9-7.

Fig. 19-2. Radio Electronics, 9/89, p. 47.

Fig. 19-3. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-65.

Fig. 19-4. Maxim, Seminar Applications Book, 1988/89, p. 149.

Fig. 19-5. Ham Radio, 5/89, p. 26.

Fig. 19-6. Popular Electronics, 10/89, p. 42.

Fig. 19-7. Maxim, Seminar Applications Book, 1988/89, p. 83.

Fig. 19-8. Reprinted from EDN, 10/2/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 19-9. Linear Technology Corp., Linear Databook, 1986, p. 8-13.

Fig. 19-10. Linear Technology Corp., Linear Databook, 1986, p. 8-43.

Fig. 19-11. Maxim, Seminar Applications Book, 1988/89, p. 78.

Fig. 19-12. Reprinted from EDN, 3/7/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 19-13. Reprinted from EDN, 8/17/89, © Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 19-14. Linear Technology Corp., Linear Databook, 1986, p. 2-112.

Fig. 19-15. Reprinted from EDN, 10/1/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 19-16. Reprinted from EDN, 3/3/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 19-17. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-62.

Chapter 20

Fig. 20-1. QST, 2/89, p. 21.

Fig. 20-2. Reprinted from EDN, 11/10/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 20-3. Hands-On Electronics / Popular Electronics, 1/89, p. 59.

Fig. 20-4. Reprinted from EDN, 3/74, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 20-5. Intersil, Component Data Catalog, 1987, p. 14-49.

Fig. 20-6. Gernsback Publications Inc., 42 New Ideas, 1984, p. 18.

Fig. 20-7. R-E Experimenters Handbook, 1987, p. 151.

Chapter 21

Fig. 21-1. Motorola, MECL System Design Handbook, 1983, p. 226.

Fig. 21-2. Reprinted from EDN, 2/78, © 1989 Cahners

Publishing Co., a division of Reed Publishing USA.

Fig. 21-3. Radio-Electronics, 2/86, p. 46.

Fig. 21-4. Reprinted from EDN, 6/83, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 21-5. Reprinted from EDN, 6/83, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 21-6. Radio-Electronics, 2/87, p. 96.

Fig. 21-7. Motorola, MECL System Design Handbook, 1983, p. 228.

Fig. 21-8. Reprinted from EDN, 5/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 21-9. NASA Tech Briefs, 8/89, p. 20.

Fig. 21-10. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-14.

Fig. 21-11. RF Design, 3/87, p. 31.

Fig. 21-12. 73 Amateur Radio, 1/89, p. 35.

Fig. 21-13. RF Design, 3/87, p. 31.

Fig. 21-14. RF Design, 3/87, p. 31.

Chapter 22

Fig. 22-1. R-E Experiments Handbook, 1989, p. 23.

Fig. 22-2. Hands-On Electronics / Popular Electronics, 1/89, p. 97.

Fig. 22-3. Reprinted from EDN, 7/21/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 22-4. Hands-On Electronics / Popular Electronics, 1/89, p. 84.

Fig. 22-5. Reprinted from EDN, 9/29/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 23

Fig. 23-1. Reprinted from EDN, 10/29/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 23-2. Reprinted from EDN, 3/4/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 23-3. Reprinted from EDN, 10/3/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 24

Fig. 24-1. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 4-303.

Chapter 25

Fig. 25-1. Reprinted from EDN, 3/21/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 25-2. Harris, Analog Product Data Book, 1988, p. 2-106.

Fig. 25-3. Teledyne Semiconductor, Data Acquisitions IC Handbook, 1985, p. 15-15.

Fig. 25-4. Siliconix, Integrated Circuits Data Book, 3/85, p.5-8.

Fig. 25-5. Signetics, RF Communications Handbook, 1989, p. 3-22.

Fig. 25-6. Reprinted from EDN, 1/20/79, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 25-7. Gernsback Publications Inc., 42 New Ideas, 1984, p. 6.

Fig. 25-8. National Semiconductor Corp., Linear Databook, 1982, p. 3-97.

Fig. 25-9. Radio-Electronics, 3/89, p. 12.

Fig. 25-10. Harris, Analog Product Data Book, 1988, p. 10-167.

Fig. 25-11. Reprinted from EDN, 9/19/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 25-12. Reprinted from EDN, 4/76, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 25-13. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-11 and 7-12.

Fig. 25-14. Radio Electronics, 4/89, p. 60.

Fig. 25-15. Reprinted from EDN, 12/8/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 25-16. Reprinted from EDN, 2/20/76, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 25-17. Reprinted from EDN, 10/17/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 26

Fig. 26-1. Siliconix, Integrated Circuits Data Book, 3/85, p. 4-15.

Fig. 26-2. Precision Monolithics Inc., 1981 Full Line Catalog, p. 6-85.

Fig. 26-3. Intersil, Applications Handbook, 1988, p. 2-38.

Fig. 26-4. Intersil, Applications Handbook, 1988, p. 3-183.

Fig. 26-5. Harris, Analog Product Data Book, 1988, p. 10-17.

Fig. 26-6. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 12.

Chapter 27

Fig. 27-1. Electronic Engineering, 12/78, p. 31.

Chapter 28

Fig. 28-1. Siliconix, Mospower Applications Handbook, p. 6-161.

Fig. 28-2. Motorola, Motorola Thyristor Device Data, Series A, 1985, p.1-5-27.

Fig. 28-3. Reprinted from EDN, 7/10/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 28-4. Siliconix, Integrated Circuits Data Book, 3/85, p. 1-11.

Chapter 29

Fig.29-1. Signetics, Fiber-Optic Communication Data and Applications, 1988, p. 5-3.

Fig. 29-2. Signetics, Fiber-Optic Communication Data and Applications, 1988, p. 3-63 and 3-64.

Fig. 29-3. Signetics, Fiber-Optic Communication Data and Applications, 1988, p. 2-3 and 2-4.

Fig. 29-4. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 5-56 and 5-58.

Fig. 29-5. Signetics, Fiber-Optic Communication Data and Applications, 1988, p. 3-3.

Fig. 29-6. Signetics, Fiber-Optic Communication Data and Applications, 1988, p. 3-7 and 3-9.

Chapter 30

Fig. 30-1. Ham Radio, 1/85, p. 51.

Fig. 30-2. Hands-On Electronics, 8/87, p. 65.

Chapter 31

Fig. 31-1. Siliconix, Integrated Circuits Data Book, 1988, p. 13-181.

Fig. 31-2. Reprinted from EDN, 9/29/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 31-3. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing,

Fig. 31-4. Harris, Analog Product Data Book, 1988, p. 10-16.

Fig. 31-5. Harris, Analog Product Data Book, 1988, p. 10-175.

Fig. 31-6. Intersil, Component Data Catalog, 1987, p. 7-45.

Fig. 31-7. Reprinted from EDN, 3/16/89, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 31-8. Raytheon, Linear and Integrated Circuits, 1989, p. 4-189.

Fig. 31-9. Popular Electronics, Fact Card No. 101.

Fig. 31-10. Popular Electronics, Fact Card No. 104.

Fig. 31-11. Popular Electronics, Fact Card No. 59.

Fig. 31-12. Popular Electronics, Fact Card No. 59.

Fig. 31-13. Popular Electronics, Fact Card No. 117.

Fig. 31-14. Popular Electronics, Fact Card No. 117.

Fig. 31-15. Popular Electronics, Fact Card No. 117.

Fig. 31-16. Popular Electronics, Fact Card No. 101.

Chapter 32

Fig. 32-1. Hands-On Electronics, 3/87, p. 97.

Fig. 32-2. Ideas for Design.

Fig. 32-3. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-4. Popular Electronics, 9/89, p. 88.

Fig. 32-5. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-6. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-7. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-8. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-9. Popular Electronics, Fact Card No. 65.

Fig. 32-10. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-17.

Fig. 32-11. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-12. Hands-On Electronics, 3/87, p. 97.

Fig. 32-13. ZeTeX, (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-153.

Fig. 32-14. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-15. Popular Electronics, 9/89, p. 23.

Fig. 32-16. Reprinted from EDN, 7/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-17. Reprinted from EDN, 9/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 32-18. Popular Electronics/Hands-On Electronics, 5/89, p. 86.

Chapter 33

Fig.33-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN5-5.

Chapter 34

34-1. National Semiconductor Corp., Linear Applications Databook, p. 1079.

Fig. 34-2. Radio-Electronics, 6/88, p. 49.
Fig. 34-3. TAB Books, The Build-It Book of Electronic Projects, p. 18.
Fig. 34-4. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 23.
Fig. 34-5. Hands-On Electronics, 9–10/ 86, p. 24.
Fig. 34-6. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.
Fig. 34-7. Reprinted from EDN, 2/21/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 35

Fig. 35-1. Signetics, RF Communications Handbook, 1989, p. 1-33.
Fig. 35-2. Signetics, RF Communications Handbook, 1989, p. 1-33.
Fig. 35-3. Intersil, Component Data Catalog, 1987, p. 7-44.
Fig. 35-4. Hands-On Electronics, Fact Card No. 29.

Chapter 36

Fig. 36-1. Electronic Engineering, 9/87, p. 27.
Fig. 36-2. Reprinted from EDN, 5/26/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 36-3. Reprinted from EDN, 2/77, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 36-4. Reprinted from EDN, 6/8/89, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 36-5. Signetics, RF Communications Handbook, 1989, p. 4-10 and 4-11.
Fig. 36-6. Reprinted from EDN, 11/24/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 37

Fig. 37-1. Reprinted from EDN, 5/83, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 38

Fig. 38-1. Reprinted from EDN, 3/21/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 38-2. Reprinted from EDN, 11/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 38-3. Reprinted from EDN, 12/13/84, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 38-4. Popular Electronics/Hands-On Electronics, 4/89, p. 23.
Fig. 38-5. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 7.
Fig. 38-6. Electronic Engineering, 9/88, p. 34.
Fig. 38-7. Electronic Engineering, 9/86, p. 34.
Fig. 38-8. Reprinted from EDN, 4/14/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 38-9. Electronic Engineering, 4/88, p. 33.
Fig. 38-10. Reprinted from EDN, 5/26/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 38-11. GE/RCA BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 8.
Fig. 38-12. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.
Fig. 38-13. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 4-312.
Fig. 38-14. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-18.
Fig. 38-15. Intersil, 1978 Databook.
Fig. 38-16. Harris, Analog Product Data Book, 1988, p. 10-109.
Fig. 38-17. Siliconix, Integrated Circuits Data Book, 1981, p. 8-51.
Fig. 38-18. Hands-On Electronics/Popular Electronics, 1/89, p. 84.
Fig. 38-19. GE/RCA BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 9.
Fig. 38-20. National Semiconductor Corp., Linear Applications Databook, p. 1118.
Fig. 38-21. Raytheon, Linear and Integrated Circuit, 1984, p. 12-8.
Fig. 38-22. Reprinted from EDN, 6/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 38-23. Maxim, Seminar Applications Book, 1988/89, p. 45.
Fig. 38-24. Raytheon, Linear and Integrated Circuits, 1984, p. 12-7.
Fig. 38-25. Harris, Analog Product Data Book, 1988, p. 10-15.
Fig. 38-26. Intersil, Component Data Catalog, 1987, p. 7-104.
Fig. 38-27. Popular Electronics, Fact Card No. 98.
Fig. 38-28. Hands-On Electronics, Fact Card No. 86.
Fig. 38-29. Harris, Analog Product Data Book, 1988, p. 10-168.
Fig. 38-30. Hands-On Electronics, Fact Card No. 86.
Fig. 38-31. Raytheon, Linear and Integrated Circuits,

1989, p. 4-188.
Fig. 38-32. GE/RCA BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 8.
Fig. 38-33. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 9.
Fig. 38-34. Hands-On Electronics, Fact Card No. 89.
Fig. 38-35. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 4-311.
Fig. 38-36. Intersil, Component Data Catalog, 1987, p. 7-44.
Fig. 38-37. Hands-On Electronics, Fact Card No. 89.

Chapter 39

Fig. 39-1. Popular Electronics/Hands-On Electronics, 3/89, p. 24.
Fig. 39-2. Popular Electronics, 6/89, p. 27.
Fig. 39-3. Elektor, 6/78, p. 6-18.

Chapter 40

Fig. 40-1. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 27.
Fig. 40-2. Linear Technology Corp., Linear Databook, 1986, p. 2-99.
Fig. 40-3. Reprinted from EDN, 8/75, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 40-4. Reprinted from EDN, 9/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 40-5. Linear Technology Design Notes, 3/89, Number 5.
Fig. 40-6. General Instrument Microelectronics, Application Note 1601, p. 3.
Fig. 40-7. Reprinted from EDN, 2/76, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 40-8. Popular Electronics, Fact Card No. 65.
Fig. 40-9. Reprinted from EDN, 10/30/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 41

Fig. 41-1. Reprinted from EDN, 7/7/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 41-2. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 12-9.
Fig. 41-3. Reprinted from EDN, 7/11/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 41-4. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 12-8.

Chapter 42

Fig. 42-1. Motorola, RF Data Manual, 1986, p. 6-226.
Fig. 42-2. Motorola, RF Data Manual, 1986, p. 6-85.
Fig. 42-3. 73 Amateur Radio, 3/89, p. 66.
Fig. 42-4. Signetics, RF Communications Handbook, 1989, p. 1-31.
Fig. 42-5. Siliconix, Small-Signal FET Data Book, 1989, p. 4-158, 4-159, and 9-42.
Fig. 42-6. R-E Experiments Handbook, 1989, p. 156.
Fig. 42-7. R-E Experiments Handbook, 1989, p. 33.
Fig. 42-8. Reprinted from EDN, 1/7/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 43

Fig. 43-1. NASA, NASA Tech Briefs, p. 55.

Chapter 44

Fig. 44-1. Elektor Electronics, 7–8/87 Supplement, p. 36.
Fig. 44-2. Reprinted from EDN, 5/74, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 44-3. Hands-On Electronics, Fact Card No. 57.
Fig. 44-4. 73 Magazine, 10/83, p. 53.
Fig. 44-5. Practical Wireless, 5/85, p. 37.

Chapter 45

Fig. 45-1. Radio-Electronics, 8/88, p. 37.
Fig. 45-2. GE, optoelectronics, Third Edition, Ch. 6, p. 115.
Fig. 45-3. Popular Electronics/Hands-On Electronics, 4/89, p. 83.
Fig. 45-4. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-155.
Fig. 45-5. Popular Electronics/Hands-On Electronics, 4/89, p. 82.
Fig. 45-6. Reprinted from EDN, 11/13/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 45-7. Hands-On Electronics, Fact Card No. 83.
Fig. 45-8. Popular Electronics, Fact Card No. 94.
Fig. 45-9. Hands-On Electronics, Fact Card No. 83.

Chapter 46

Fig. 46-1. Linear Technology, Application Note 9, p. 6.
Fig. 46-2. Maxim, Maxim Advantage, p. 45.
Fig. 46-3. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-2.
Fig. 46-4. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 19.

Fig. 46-5. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 15.

Fig. 46-6. Harris, Analog Product Data Book, 1988, p. 10-181.

Fig. 46-7. Maxim, Maxim Advantage, p. 45.

Fig. 46-8. Siliconix, Integrated Circuits Data Book, 1988, p. 5-172.

Chapter 47

Fig. 47-1. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-64.

Fig. 47-2. Linear Technology Corp., Linear Databook, 1986, p. 8-40.

Chapter 48

Fig. 48-1. Fairchild Camera & Instrument Corp., Linear Databook, 1982, p. 4-88.

Fig. 48-2. National Semiconductor Corp., Linear Databook, 1982, p. 10-63.

Fig. 48-3. Hands-On Electronics/Popular Electronics, 1/89, p. 39.

Fig. 48-4. Hands-On Electronics/Popular Electronics, 11/88, p. 39.

Fig. 48-5. Popular Electronics, Fact Card No. 59.

Chapter 49

Fig. 49-1. R-E Experimenters Handbook, 1987, p. 129.

Fig. 49-2. Popular Electronics, 6/89, p. 25.

Fig. 49-3. Maxim, Seminar Applications Book, 1988/89, p. 81.

Fig. 49-4. Motorola, Motorola TMOS Power FET Design Ideas, p. 35.

Fig. 49-5. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-14.

Fig. 49-6. Popular Electronics, 10/89, p.26.

Chapter 50

Fig. 50-1. Elektor Electronics, 7 – 8/87 Supplement, p. 55.

Fig. 50-2. Reprinted from EDN, 10/27/88, © Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 50-3. Hands-On Electronics/Popular Electronics, 1/89, p. 27.

Fig. 50-4. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 22.

Fig. 50-5. Motorola, Motorola Thyristor Device Data, Series A, 1985, p. 1-6-54.

Fig. 50-6. Hands-On Electronics, Fact Card No. 37.

Fig. 50-7. Motorola, Motorola Thyristor Device Data, Series A, 1985, p. 1-6-48.

Fig. 50-8. Modern Electronics, 11/85, p. 44.

Fig. 50-9. Motorola, Motorola Thyristor Device Data, Series A, 1985, p. 1-6-10.

Fig. 50-10. GE, Optoelectronics, Third Edition, Ch. 6, p. 106.

Fig. 50-11. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-158.

Fig. 50-12. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.

Fig. 50-13. GE, Optoelectronics, Third Edition, Ch. 6, p. 110.

Chapter 51

Fig. 51-1. TAB Books, Build Your Own Laser, Phaser, Ion Ray Gun, 1983, p. 13.

Fig. 51-2. Siliconix, Mospower Applications Handbook, p. 6-184.

Chapter 52

Fig. 52-1. GE, Optoelectronics, Third Edition, Ch. 6, p. 109.

Fig. 52-2. GE, Optoelectronics, Third Edition, Ch. 6, p. 112.

Fig. 52-3. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 19.

Fig. 52-4. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 3-3 and 3-4.

Fig. 52-5. Reprinted from EDN, August 1976, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 52-6. Harris, Analog Product Data Book, 1988, p. 10-182.

Fig. 52-7. GE, Optoelectronics, Third Edition, Ch. 6, p. 109.

Fig. 52-8. GE, Optoelectronics, Third Edition, Ch. 6, p. 107.

Fig. 52-9. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 7.

Fig. 52-10. Hands-On Electronics, Fact Card No. 57.

Fig. 52-11. GE, Optoelectronics, Third Edition, Ch. 6, p. 108.

Fig. 52-12. GE, Application Note 200.35, p. 15.

Chapter 53

Fig. 53-1. Hands-On Electronics, 9/87, p. 96.

Fig. 53-2. Popular Electronics, Fact Card No. 113.

Fig. 53.3 Harris, Analog Product Data Book, 1988, p. 10-54.

Chapter 54

Fig. 54-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 5-350 and 5-352.

Chapter 55

Fig. 55-1. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.
Fig. 55-2. Hands-On Electronics, Fact Card No. 29.
Fig. 55-3. Hands-On Electronics, Fact Card No. 29.

Chapter 56

Fig. 56-1. Reprinted from EDN, 5/2/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 56-2. Popular Electronics, 10/89, p. 104.
Fig. 56-3. Hands-On Electronics, Winter 1985, p. 31.
Fig. 56-4. Ham Radio, 4/86, p. 24.
Fig. 56-5. Gernsback Publications Inc., 42 New Ideas, 1984, p. 24.
Fig. 56-6. Ham Radio, 12/88, p. 19.
Fig. 56-7. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 12-5.
Fig. 56-8. Reprinted from EDN, 3/23/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 56-9. 73 Amateur Radio, 6/89, p. 44.
Fig. 56-10. Hands-On Electronics, 2/87, p. 92.
Fig. 56-11. Elektor Electronics, 7–8/87 Supplement, p. 23.
Fig. 56-12. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 20.
Fig. 56-13. GE, Optoelectronics, Third Edition, Ch. 6, p. 111.
Fig. 56-14. Hands-On Electronics, 7–8/86, p. 51.
Fig. 56-15. Intersil, Component Data Catalog, 1987, p. 14-70.
Fig. 56-16. Intersil, Component Data Catalog, 1987, p. 14-121.
Fig. 56-17. NASA, NASA Tech Briefs, 1/89, p. 19.
Fig. 56-18. Reprinted from EDN, 5/83, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 56-19. 73 Amateur Radio, 8/88, p. 24.
Fig. 56-20. Linear Technology Corp., Linear Databook, 1986, p. 2-96.
Fig. 56-21. Radio-Electronics, 9/87, p. 32.
Fig. 56-22. Hands-On Electronics/Popular Electronics, 12/88, p. 61.
Fig. 56-23. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-137.
Fig. 56-24. GE/RCA BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 12.
Fig. 56-25. Electronic Engineering, 11/86, p. 34.
Fig. 56-26. Popular Electronics, 10/89, p. 84.
Fig. 56-27. Teledyne Semiconductor, Data Acquisition IC Handbook, 1985, p. 15-11.
Fig. 56-28. GE, Optoelectronics, Third Edition, Ch. 6, p. 112.
Fig. 56-29. Reprinted from EDN, 3/79, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 57

Fig. 57-1. Reprinted from EDN, 3/7/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 57-2. Electronic Engineering, Applied Ideas, 5/88, p. 25.
Fig. 57-3. Signetics. 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-63.
Fig. 57-4. Reprinted from EDN, 9/15/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 58

Fig. 58-1. Radio-Electronics, 12/85, p. 38.

Chapter 59

Fig. 59-1. Popular Electronics/Hands-On Electronics, 3/89, p. 82.
Fig. 59-2. GE, Application Note 90.16, p. 25.
Fig. 59-3. Popular Electronics/Hands-On Electronics, 3/89, p. 83.
Fig. 59-4. Popular Electronics/Hands-On Electronics, 4/89, p. 22.
Fig. 59-5. Popular Electronics/Hands-On Electronics, 3/89, p. 82.
Fig. 59-6. Radio-Electronics, 10/88, p. 51.
Fig. 59-7. Radio-Electronics, 4/85, p. 40.
Fig. 59-8. Siliconix, Small-Signal FET Data Book, 1989, p. 6-69 and 6-70.
Fig. 59-9. R-E Experimenter's Handbook, 1989, p. 41.
Fig. 59-10. RF Design, 7/89, p. 53.
Fig. 59-11. GE, Application Note 200.85, p. 16.
Fig. 59-12. GE, Application Note 200.85, p. 19.
Fig. 59-13. Harris, Analog Product Data Book, 1988, p. 10-109.
Fig. 59-14. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN13-24.
Fig. 59-15. Gernsback Publications Inc., 44 New Ideas, 1985, p. 2.
Fig. 59-16. Radio-Electronics, 6/89, p. 42.
Fig. 59-17. Hands-On Electronics, 3/87, p. 26.

Chapter 60

Fig. 61-1. National Semiconductor Corp., Linear Applications Databook, p. 1028.

Fig. 60-2. Popular Electronics, 7/89, p. 23.
Fig. 60-3. Harris, Analog Product Data Book, 1988, p. 10-170.
Fig. 60-4. Popular Electronics, Fact Card No. 107.
Fig. 60-5. Harris, Analog Product Data Book, 1988, p. 10-176.

Chapter 61

Fig. 61-1. Reprinted from EDN, 11/28/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 61-2. Signetics, 1987 Linear Data Manual Vol.2: Industrial, 10/10/86, p. 4-260.
Fig. 61-3. Electronic Design, 11/24/88, p. 222.
Fig. 61-4. Reprinted from EDN, 11/28/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 61-5. Popular Electronics, 10/89, p. 23.
Fig. 61-6. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-59.
Fig. 61-7. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/78, p. 7-59.
Fig. 61-8. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 4-3.
Fig. 61-9. Signetics, RF Communications Handbook, 1989, p. 4-9.

Chapter 62

Fig. 62-1. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 53.
Fig. 62-2. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 29.
Fig. 62-3. Reprinted with permission from Electronic Design. Copyright 1988, Penton Publishing.
Fig. 62-4. Reprinted from EDN, 8/20/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 62-5. Reprinted from EDN, 3/77, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 62-6. Signetics, 1987 Linear Data Manual Vol.2: Industrial, 2/87, p. 7-63.
Fig. 62-7. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN1-7.
Fig. 62-8. NASA, NASA Tech Briefs, 2/89, p. 26.
Fig. 62-9. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN11-7.
Fig. 62-10. Maxim, Maxim Advantage, p. 47.
Fig. 62-11. Signetics, 1987 Linear Data Manual Vol.2: Industrial, 2/87, p. 8-94 and 8-95.
Fig. 62-12. Fairchild Camera and Instrument Corp., Linear Databook, 1982, p. 4-89.
Fig. 62-13. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 10-22.
Fig. 62-14. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-22.
Fig. 62-15. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-162.
Fig. 62-16. Linear Technology Corp., Linear Databook, 1986, p. 5-15.

Chapter 63

Fig. 63-1. Siliconix, Integrated Circuits Data Book, 1988, p. 13-203.
Fig. 63-2. Maxim, Seminar Applications Book, 1988/89, p. 61.
Fig. 63-3. Intersil, Applications Handbook, 1988, p. 5-6.
Fig. 63-4. Siliconix, Integrated Circuits Data Book, 1988, p. 13-200.
Fig. 63-5. Harris, Analog Product Data Book, 1988, p. 10-23.
Fig. 63-6. Intersil, Applications Handbook, 1988, p. 5-6.
Fig. 63-7. Harris, Analog Product Data Book, 1988, p. 10-12.

Chapter 64

Fig. 64-1. Signetics, Analog Data Manual, 1983, p. 4-26.
Fig. 64-2. Signetics, Analog Data Manual, 1983, p. 4-32.
Fig. 64-3. Signetics, Analog Data Manual, 1983, p. 4-33.

Chapter 65

Fig. 65-1. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 3-10 and 3-11.
Fig. 65-2. National Semiconductor Corp., Linear Applications, p. 1083.

Chapter 66

Fig. 66-1. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-211.

Chapter 67

Fig. 67-1. GE, Optoelectronics, Third Edition, Ch. 6, p. 141.
Fig. 67-2. GE, Optoelectronics, Third Edition, Ch. 6, p. 135.
Fig. 67-3. GE, Optoelectronics, Third Edition, Ch. 6, p. 144.
Fig. 67-4. GE, Optoelectronics, Third Edition, Ch. 6, p. 134.
Fig. 67-5. GE, Optoelectronics, Third Edition, Ch. 6, p. 147.

Fig. 67-6. GE, Optoelectronics, Third Edition, Ch. 6, p. 127.
Fig. 67-7. Reprinted from EDN, 1/82, © 1989 Cahner Publishing Co., a division of Reed Publishing USA.
Fig. 67-8. GE, Optoelectronics, Third Edition, Ch. 6, p. 123.
Fig. 67-9. GE, Optoelectronics, Third Edition, Ch. 6, p. 126.
Fig. 67-10. GE, Optoelectronics, Third Edition, Ch. 6, p. 108.
Fig. 67-11. GE, Optoelectronics, Third Edition, Ch. 6, p. 122.
Fig. 67-12. GE, Optoelectronics, Third Edition, Ch. 6, p. 135.
Fig. 67-13. GE, Optoelectronics, Third Edition, Ch. 6, p. 145.
Fig. 67-14. GE, Optoelectronics, Third Edition, Ch. 6, p. 138.
Fig. 67-15. Reprinted from EDN, 11/10/88, © 1989 Cahner Publishing Co., a division of Reed Publishing USA.

Chapter 68

Fig. 68-1. Reprinted from EDN, 8/7/86, © 1989 Cahner Publishing Co., a division of Reed Publishing USA.
Fig. 68-2. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-17.
Fig. 68-3. GE, Application Note 90.16, p. 27.
Fig. 68-4. Elektor Electronics, 7–8/87 Supplement, p. 36.
Fig. 68-5. Harris, Analog Product Data Book, 1988, p. 10-15.
Fig. 68-6. Elektor Electronics, 7–8/87 Supplement, p. 63.
Fig. 68-7. Reprinted from EDN, 1/79, © 1989 Cahner Publishing Co., a division of Reed Publishing USA.
Fig. 68-8. Elektor Electronics, 7–8/87 Supplement, p. 59.
Fig. 68-9. Reprinted from EDN, 7/74, © 1989 Cahners Publishing Co., A division of Reed Publishing USA.
Fig. 68-10. Harris, Analog Product Data Book, 1988, p. 10-97.
Fig. 68-11. Reprinted from EDN, 1/77, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 68-12. Radio-Electronics, 8/86, p. 83.
Fig. 68-13. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN20-11.
Fig. 68-14. Fairchild Camera and Instrument, Linear Databook, 1982, p. 4-71.
Fig. 68-15. Reprinted from EDN, 8/76, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 68-16. Reprinted from EDN, 5/11/89, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 68-17. Reprinted from EDN, 4/82, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 68-18. Hands-On Electronics, Fact Card No. 49.
Fig. 68-19. Reprinted from EDN, 2/2/89 © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 68-20. Hands-On Electronics/Popular Electronics, 1/89, p. 26.
Fig. 68-21. GE/RCA, BiMOS Operational Amplifier Circuit Ideas, 1987, p. 9.
Fig. 68-22. Reprinted from EDN, 1/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 68-23. Harris, Analog Product Data Book, 1988, p. 10-108.

Chapter 69

Fig. 69-1. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-62.
Fig. 69-2. Elektor Electronics, 7-8/87 Supplement, p. 50.
Fig. 69-3. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-13.
Fig. 69-4. Reprinted from EDN, 3/82, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 69-5. Radio-Electronics, 10/84, p. 32.
Fig. 69-6. 73 Amateur Radio, 8/88, p. 47.
Fig. 69-7. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 4-2.
Fig. 69-8. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-62.
Fig. 69-9. Ham Radio, 12/88, p. 26.

Chapter 70

Fig. 70-1. Harris, Analog Product Data Book, 1988, p. 10-14.
Fig. 70-2. Reprinted from EDN, 8/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 70-3. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 4-67 and 4-68.

Chapter 71

Fig. 71-1. Popular Electronics/Hands-On Electronics, 4/89, p. 26.
Fig. 71-2. Hands-On Electronics, 1/87, p. 101.
Fig. 71-3. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-14.
Fig. 71-4. Hands-On Electronics, 11/87, p. 32.

Fig. 71-5. GE, Optoelectronics, Third Edition, Ch. 6, p. 105.
Fig. 71-6. Popular Electronics, 9/89, p. 22.
Fig. 71-7. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 24.

Chapter 72

Fig. 72-1. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 1.
Fig. 72-2. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 25.
Fig. 72-3. Popular Electronics.
Fig. 72-4. Motorola, Motorola Power FET Design Ideas, 1985, p. 2.
Fig. 72-5. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 3-14.
Fig. 72-6. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/6/86, p. 7-246.
Fig. 72-7. Hands-On Electronics, 7-8/86, p. 43.
Fig. 72-8. Siliconix, Mospower Applications Handbook, p. 6-180.

Chapter 73

Fig. 73-1. GE, Optoelectronics, Third Edition, Ch. 6, p. 152.
Fig. 73-2. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 38.
Fig. 73-3. 73 Amateur Radio, 6/89, p. 61.
Fig. 73-4. Texas Instruments, Linear and Interface Circuits Applications, 1987, p. 12-5.
Fig. 73-5. QST, 6/88, p. 48.
Fig. 73-6. Texas Instruments, Linear Circuits Data Book, 1989, p. 2-74.
Fig. 73-7. Maxim, 1986 Power Supply Circuits, p. 27.
Fig. 73-8. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 6-36.
Fig. 73-9. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 6-36.
Fig. 73-10. Reprinted from EDN, 5/5/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 73-11. Hands-On Electronics, 7-8/86.
Fig. 73-12. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 6-34.
Fig. 73-13. 73 Amateur Radio, 5/88, p. 90.
Fig. 73-14. Reprinted from EDN, 7/7/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 73-15. Harris, Analog Product Data Book, 1988, p. 10-16.
Fig. 73-16. Hands-On Electronics, Fact Card No. 49.
Fig. 73-17. Popular Electronics, Fact Card No. 100.
Fig. 73-18. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-13.
Fig. 73-19. Harris, Analog Product Data Book, 1988, p. 10-107.
Fig. 73-20. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 12.
Fig. 73-21. Reprinted from EDN, 4/20/77, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 73-22. Intersil, Applications Handbook, 1988, p. 6-17.
Fig. 73-23. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 14.
Fig. 73-24. Electronic Engineering, 2/87, p. 44.
Fig. 73-25. Popular Electronics, Fact Card No. 100.
Fig. 73-26. Intersil, Applications Handbook, 1988, p. 6-70.
Fig. 73-27. Reprinted with permission from Electronic Design. Copyright 1988, Penton Publishing.
Fig. 73-28. Maxim, Seminar Applications Book, 1988/89, p. 83.
Fig. 73-29. Intersil, Applications Handbook, 1988, p. 6-18.
Fig. 73-30. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-15.
Fig. 73-31. Popular Electronics, Fact Card No. 100.
Fig. 73-32. Popular Electronics, Fact Card No. 95.
Fig. 73-33. Maxim, 1986 power Supply Circuits, p. 26.

Chapter 74

Fig. 74-1. NASA, NASA Tech Briefs, 1/88, p. 22.
Fig. 74-2. Reprinted from EDN 6/22/89, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 74-3. Motorola, Motorola, TMOS Power FET Design Ideas, 1985, p. 40.
Fig. 74-4. Popular Electronics, 10/89, p. 37.
Fig. 74-5. Hands-On Electronics, 12/86, p. 93.
Fig. 74-6. Electronic Engineering, Applied Ideas, 3/86, p. 33.
Fig. 74-7. Reprinted from EDN, 9/15/82 and 1/6/83, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 79-8. Reprinted from EDN, 10/17/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 74-9. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN2-7.
Fig. 74-10. NASA, NASA Tech Briefs, Summer 1985, p. 48.

Fig. 74-11. Hands-On Electronics, 12/86, p. 92.

Chapter 75

Fig. 75-1. Siliconix, Mospower Applications Handbook, p. 6-62.
Fig. 75-2. 73 Amateur Radio, 4/88, p. 20.
Fig. 75-3. Reprinted from EDN, 12/8/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 75-4. Linear Technology Corp., Linear Application Handbook, 1987, p. AN8-4.
Fig. 75-5. Ham Radio, 7/89, p. 20.
Fig. 75-6. Popular Electronics, Fact Card No. 100.
Fig. 75-7. Popular Electronics, Fact Card No. 95.

Chapter 76

Fig. 76-1. Hands-On Electronics, 1-2/86, p. 96.
Fig. 76-2. Maxim, 1986 Power Supply Circuits, p. 120.

Chapter 77

Fig. 77-1. Hands-On Electronics, 12/87, p. 73.
Fig. 77-2. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN9-9.
Fig. 77-3. Microwaves and RF, 3/86, p. 143.
Fig. 77-4. Reprinted from EDN, 6/72, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 77-5. Reprinted from EDN, 7/7/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 77-6. Reprinted from EDN, 10/16/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 77-7. Reprinted from EDN, 3/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 77-8. Harris, Analog Product Data Book, 1988, p. 10-182.
Fig. 77-9. Linear Technology Corp., Application Note 9, p. 9.
Fig. 77-10. Hands-On Electronics, 9/87, p. 96.

Chapter 78

Fig. 78-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-5.
Fig. 78-2. Siliconix, Integrated Circuits Data Book, 1988, p. 5-128.
Fig. 78-3. Intersil, Applications Handbook, 1988, p. 2-34.
Fig. 78-4. Harris, Analog Product Data Book, 1988, p. 10-13.
Fig. 78-5. Harris, Analog Product Data Book, 1988, p. 10-13.
Fig. 78-6. Harris, Analog Product Data Book, 1988, p. 10-169.

Chapter 79

Fig. 79-1. 73 Magazine, 7/77, p. 35.
Fig. 79-2. Reprinted from EDN, 4/83, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 79-3. TAB Books, 104 Weekend Electronics Projects, p. 70.
Fig. 79-4. Contributed by William Sheets.
Fig. 79-5. Reprinted from EDN, 6/12/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 79-6. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 16.

Chapter 80

Fig. 80-1. Hands-On Electronics, 7/87, p. 47.
Fig. 80-2. RF Design, 12/86, p. 41.
Fig. 80-3. Popular Electronics/Hands-On Electronics, 5/89, p. 85.
Fig. 80-4. Hands-On Electronics, Fact Card No. 57.
Fig. 80-5. Siliconix, Small-Signal FET Data Book, 1/86, p. 7-28.

Chapter 81

Fig. 81-1. Popular Electronics, 9/89, p. 41.
Fig. 81-2. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 15.
Fig. 81-3. Reprinted from EDN, 11/10/88, © Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 81-4. Reprinted from EDN, 5/82, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 81-5. Reprinted from EDN, 11/27/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 81-6. Linear Technology Corp., Linear Databook, 1986, p. 5-33.

Chapter 82

Fig. 82-1. Reprinted from EDN, 11/13/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 82-2. Harris, Analog Product Data Book, 1988, p. 10-14.

Chapter 83

Fig. 83-1. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/14/86, p. 7-33.
Fig. 83-2. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 11/14/86, p. 4-115.
Fig. 83-3. Radio-Electronics, 8/87, p. 39.
Fig. 83-4. Signetics, RF Communications Handbook,

1989, p. 3-80 and 3-81.

Fig. 83-5. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-66 and 7-67.

Fig. 83-6. RF Design, 3/87, p. 46.

Fig. 83-7. 73 Amateur Radio, 10.86, p. 54.

Fig. 83-8. GE, Optoelectronics, Third Edition, Ch. 6, p. 128.

Fig. 83-9. Hands-On Electronics/Popular Electronics, 1/89, p. 35.

Fig. 83-10. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 7-25.

Chapter 84

Fig. 84-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/6/86, p. 4-137.

Fig. 84-2. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 17.

Chapter 85

Fig. 85-1. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-66 and 7-67.

Fig. 85-2. Hands-On Electronics/Popular Electronics, 12/88, p. 24.

Fig. 85-3. National Semiconductor Corp., Voltage Regulator Handbook, p. 10-59.

Chapter 86

Fig. 86-1. QST, 9/85, p. 41.

Fig. 86-2. Motorola, RF Data Manual, 1986, p. 6-221.

Fig. 86-3. Microwaves and RF, 9/85, p. 191.

Fig. 86-4. Teledyne Semiconductor, Data and Design Manual, 1981, p. 11-178.

Fig. 86-5. Harris, Analog Product Data Book, 1988, p. 10-58.

Fig. 86-6. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 20.

Fig. 86-7. Motorola, RF Data Manual, 1986, p. 6-182.

Fig. 86-8. Radio-Electronics, 3/82, p. 59.

Fig. 86-9. Motorola, RF Data Manual, 1986, p. 6-236.

Chapter 87

Fig. 87-1. Intersil, Component Data Catalog, p. 7-96.

Fig. 87-2. Harris, Analog Product Data Book, 1988, p. 10-22.

Fig. 87-3. Harris, Analog Product Data Book, 1988, p. 10-54.

Fig. 87-4. Harris, Analog Product Data Book, 1988, p. 10-22.

Fig. 87-5. Harris, Analog Product Data Book, 1988, p. 10-37.

Fig. 87-6. Siliconix, Integrated Circuits Data Book, 1988, 5-127.

Fig. 87-7. Harris, Analog Product Data Book, 1988, p. 10-14.

Fig. 87-8. Harris, Analog Product Data Book, 1988, p. 10-22.

Fig. 87-9. GE/RCA, BiMOS Operational Amplifier Circuit Ideas, 1987, p. 13.

Chapter 88

Fig. 88-1. Popular Electronics/Hands-On Electronics, 3/89, p. 24.

Fig. 88-2. Popular Electronics/Hands-On Electronics, 12/88, p. 26.

Chapter 89

Fig. 89-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN5-7.

Fig. 89-2. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 7.

Fig. 89-3. Linear Technology Corp., Linear Databook Supplement, 1988, p. S2-34.

Fig. 89-4. Hands-On Electronics, 9/87, p. 97.

Chapter 90

Fig. 90-1. Popular Electronics, 10/89, p. 84.

Fig. 90-2. National Semiconductor, Linear Databook, 1982, p. 3-289.

Fig. 90-3. Hands-On Electronics, Spring 1985, p. 35.

Fig. 90-4. National Semiconductor Corp., Audio/Radio Handbook, 1980, p. 4-29.

Fig. 90-5. Hands-On Electronics, 5-6/86, p. 86.

Fig. 90-6. Reprinted from EDN, 8/4/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 90-7. Everyday Electronics, 5/88, p. 292.

Fig. 90-8. Hands-On Electronics, Winter 1985, p. 72.

Fig. 90-9. Radio-Electronics, 12/81, p. 53.

Fig. 90-10. Reprinted with permission from Electronic Design. Copyright 1980, Penton Publishing.

Fig. 90-11. National Semiconductor Corp., Audio/Radio Handbook, 1980, p. 4-39.

Fig. 90-12. National Semiconductor Corp., Linear Databook, 1982, p. 3-289.

Fig. 90-13. Texas Instruments, Complex Sound Generator Bulletin No. DL-S 12612, p. 15.

Chapter 91

Fig. 91-1. GE, Optoelectronics, Third Edition, Ch. 6, p. 131.

Chapter 92

Fig. 92-1. Reprinted from EDN, 4/17/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 92-2. ZeTeX (formerly Ferranti), Technical Handbook Super E-Line Transistors, 1987, p. SE-154.

Fig. 92-3. Reprinted from EDN, 12/8/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 93

Fig. 93-1. Gernsback Publications Inc., 42 New Ideas, 1984, p. 4.

Fig. 93-2. Hands-On Electronics, 11/87, p. 92.

Fig. 93-3. Radio-Electronics, 12/81, p. 54.

Fig. 93-4. Radio-Electronics, 12/81, p. 53.

Fig. 93-5. Texas Instruments, Complex Sound Generator Bulletin No. DL-S 12612, p. 11.

Chapter 94

Fig. 94-1. Reprinted from EDN, 8/5/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 94-2. Hands-On Electronics, 3/87, p. 26.

Chapter 95

Fig. 95-1. Harris, Analog Product Data Book, 1988, p. 10-97.

Fig. 95-2. Siliconix, Integrated Circuits Data Book, 3/85, p. 5-8.

Chapter 96

Fig. 96-1. Reprinted from EDN, 4/30/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 96-2. Reprinted from EDN, 5/28/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 96-3. Exar, Telecommunications Databook, 1986, p. 9-24.

Fig. 96-4. Exar, Telecommunications Databook, 1986, p. 9-24.

Chapter 97

Fig. 97-1. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-28.

Fig. 97-2. Intersil, Component Data Catalog, 1987, p. 7-96.

Chapter 98

Fig. 98-1. Hands-On Electronics, 2/87, p. 65.

Chapter 99

Fig. 99-1. Motorola, TMOS Power FET Design Ideas, 1985, p. 48.

Fig. 99-2. Reprinted from EDN, 2/18/88, Cahner Publishing Co., a division of Reed Publishing USA.

Fig. 99-3. Reprinted from EDN, 5/79, © 1989 Cahner Publishing Co., a division of Reed Publishing USA.

Fig. 99-4. Siliconix, Mospower Applications Handbook, p. 6-185.

Fig. 99-5. Reprinted from EDN, 6/76, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 100

Fig. 100-1. Reprinted from EDN, 3/3/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 100-2. Reprinted from EDN, 5/26, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 100-3. Electronic Engineering, Applied Ideas, 11/88, p. 28.

Chapter 101

Fig. 101-1. Hands-On Electronics, 11/87, p. 84.

Fig. 101-2. Popular Electronics/Hands-On Electronics, 5/89, p. 87.

Fig. 101-3. Hands-On Electronics, 11/87, p. 85.

Chapter 102

Fig. 102-1. Exar, Telecommunications Databook, 1986, p. 4-2.

Fig. 102-2. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-23.

Fig. 102-3. Hands-On Electronics, Winter 1985, p. 49.

Fig. 102-4. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 12/2/86, p. 6-23.

Fig. 102-5. Popular Electronics, 9/89, p. 27.

Fig. 102-6. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S13-15.

Fig. 102-7. Popular Electronics, Fact Card No. 104.

Fig. 102-8. Exar, Telecommunications Databook, 1986, p. 11-61.

Fig. 102-9. Siliconix, Integrated Circuits Data Book, 3/85, p. 7-12.

Fig. 102-10. 73 Magazine, 10.78, p. 78.

Fig. 102-11. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S13-16.

Fig. 102-12. Reprinted from EDN, 8/78, © 1989

Cahners Publishing Co., a division of Reed Publishing. USA.

Fig. 102-13. Siliconix, Integrated Circuits Data Book, 3/85, p. 7-11.

Fig. 102-14. Reprinted from EDN, 10/20/79, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 102-15. Intersil, Databook 1987, p. 7-8.

Fig. 102-16. GE, Optoelectronics, Third Edition, Ch. 6, p. 128.

Fig. 102-17. Reprinted from EDN, 8/79, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 102-18. Radio-Electronics, 10/83, p. 56.

Fig. 102-19. Electronic Engineering, 2/87, p. 40.

Fig. 102-20. Reprinted from EDN, 4/27/89, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 102-21. Hands-On Electronics, 10/87, p. 95.

Fig. 102-22. Hands-On Electronics, 9-10/86, p. 88.

Fig. 102-23. Popular Electronics/Hands-On Electronics, 5/89, p. 86.

Fig. 102-24. GE, Optoelectronics, Third Edition, Ch. 6, p. 124.

Fig. 102-25. Hands-On Electronics, 7-8/86, p. 87.

Fig. 102-26. TAB Books, 101 Sound, Light, and Power IC Projects.

Fig. 102-27. Hands-On Electronics, 9-10/86, p. 105.

Chapter 103

Fig. 103-1. Motorola, Motorola Thyristor Device Data, Series A 1985, p. 1-6-60.

Fig. 103-2. Linear Technology Corp., Application Note 22, p. 10.

Fig. 103-3. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN5-1.

Fig. 103-4. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 24.

Fig. 103-5. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN4-7.

Fig. 103-6. Radio-Electronics, 5/85, p. 110.

Chapter 104

Fig. 104-1. Siliconix, Integrated Circuits Data Book, 1988, p. 13-204.

Fig. 104-2. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/14/86, p. 5-58.

Fig. 104-3. Linear Technology Corp., Linear Databook, 1986, p. 8-43.

Chapter 105

Fig. 105-1. Electronic Engineering, Applied Ideas, 11/88, p. 28.

Chapter 106

Fig. 106-1. Gernsback Publications Inc., 42 New Ideas, 1984, p. 18.

Fig. 106-2. Popular Electronics, 8/89, p. 29.

Chapter 107

Fig. 107-1. TAB Books, 44 Electronics Projects for the Darkroom.

Fig. 107-2. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/14/86, p. 5-58.

Fig. 107-3. Raytheon, Linear and Integrated Circuits, 1989, p. 8-16.

Fig. 107-4. R-E Experimenters Handbook, 1987, p. 11.

Fig. 107-5. NASA, NASA Tech Briefs, 10/87, p. 34.

Fig. 107-6. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN3-14.

Fig. 107-7. Hands-On Electronics, 9-10/86, p. 32.

Chapter 108

Fig. 108-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN7-12.

Chapter 109

Fig. 109-1. Radio-Electronics, 7/85, p. 16.

Fig. 109-2. GE, Semiconductor Data Handbook, Second Edition, p. 905.

Fig. 109-3. Motorola, Motorola Thyrister Device Data, Series A 1985, p. 1-6-43.

Chapter 110

Fig. 110-1. Reprinted from EDN, 9/13/86, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 110-2. Intersil, Databook 1987, p. 7-101.

Fig. 110-3. Radio-Electronics, 5/87, p. 129.

Fig. 110-4. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 7-13.

Fig. 110-5. TAB Books, The Build-It Book of Electronic Projects, p. 32.

Chapter 111

Fig. 111-1. Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 3-11 and 3-12.

Fig. 111-2. National Semiconductor Corp., Audio/Radio Handbook, 1980, p. 2-61.

Fig. 111-3. Popular Electronics, Fact Card No. 113.
Fig. 111-4. Reprinted from EDN, 7/7/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 111-5. Signetics, Analog Data Manual, 1983, p. 10-93.

Chapter 112

Fig. 112-1. Popular Electronics, 9/89, p. 88.
Fig. 112-2. Popular Electronics, 9/89, p. 88.
Fig. 112-3. Hands-On Electronics, 9/87, p. 89.
Fig. 112-4. Hands-On Electronics, 9/87, p. 89.
Fig. 112-5. National Semiconductor Corp., Transistor Databook, 1982, p. 7-11.
Fig. 112-6. Popular Electronics, 7/89, p. 25.
Fig. 112-7. Contributed by William Sheets.
Fig. 112-8. Contributed by William Sheets.
Fig. 112-9. Hands-On Electronics, Fact Card No. 86.

Chapter 113

Fig. 113-1. Linear Technology Corp., 1986 Linear Databook, p. 2-69.
Fig. 113-2. Fairchild Camera and Instrument Corp., Linear Databook, 1982, p. 4-71.
Fig. 113-3. Electronics Today International, 10/75, p. 64.

Chapter 114

Fig. 114-1. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 18.
Fig. 114-2. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 21.
Fig. 114-3. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 19.
Fig. 114-4. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 7-173.
Fig. 114-5. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 7-173.
Fig. 114-6. Radio-Electronics, 3/86, p. 8.
Fig. 114-7. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 7-173.
Fig. 114-8. Signetics, 1987 Linear Data Manual Vol. 1: Communications, 2/87, p. 7-173.

Chapter 115

Fig. 115-1. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 44.
Fig. 115-2. Radio-Electronics, 6/89, p. 45.
Fig. 115-3. TAB Books, Build Your Own Laser, Phaser, Ion Ray Gun, 1983, p. 292.
Fig. 115-4. 73 Amateur Radio, 1/87, p. 46.
Fig. 115-5. GE, Optoelectronics, Third Edition, Ch. 6, p. 118.
Fig. 115-6. Hands-On Electronics, 11/87, p. 44.
Fig. 115-7. R-E Experimenters Handbook, 1987, p. 161.
Fig. 115-8. Radio-Electronics, 9/89, p. 43.
Fig. 115-9. Popular Electronics, 8/89, p. 27.
Fig. 115-10. 73 Amateur Radio, 3/88, p. 96.
Fig. 115-11. Ham Radio, 9/89, p. 56.
Fig. 115-12. GI Microelectronics Gimini, TV Games Section 4B, p. 22.
Fig. 115-13. Radio-Electronics, 3/89, p. 43.
Fig. 115-14. QST, 3/86, p. 32.
Fig. 115-15. 73 Amateur Radio, 1/87, p. 48.
Fig. 115-16. Radio-Electronics, 3/88, p. 54.
Fig. 115-17. 73 Amateur Radio, 6/89, p. 20.
Fig. 115-18. Contributed by William Sheets.

Chapter 116

Fig. 116-1. TAB Books, 101 Sound, Light, and Power IC Projects.
Fig. 116-2. National Semiconductor Corp., Audio/Radio Handbook, 1980, p. 5-11.
Fig. 116-3. Electronic Engineering, 1/79, p. 27.
Fig. 116-4. Radio-Electronics, 9/85, p. 96.

Chapter 117

Fig. 117-1. Reprinted from EDN, 4/3/86, ©1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 117-2. Popular Electronics, 9/89, p. 102.
Fig. 117-3. R-E Experimenters Handbook, p. 92.
Fig. 117-4. GE, Optoelectronics, Third Edition, Ch. 6, p. 149.
Fig. 117-5. National Semiconductor Corp., Linear Applications Databook, p. 1083.
Fig. 117-6. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 26.
Fig. 117-7. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 25.
Fig. 117-8. Reprinted with permission from Electronic Design. Copyright 1975, Penton Publishing.
Fig. 117-9. Popular Electronics/Hands-On Electronics, 2/89, p. 46.
Fig. 117-10. TAB Books, Build Your Own Laser, Phaser, Ion Ray Gun, 1983, p. 170.
Fig. 117-11. Electronics Today International, 2/82, p. 89.

Chapter 118

Fig. 118-1. GE, Application Note 90.88, p. 7.

Fig. 118-2. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN4-3.

Fig. 118-3. Harris, Analog Product Data Book, 1988, p. 10-149.

Fig. 118-4. Harris, Analog Product Data Book, 1988, p. 10-58.

Fig. 118-5. Harris, Analog Product Data Book, 1988, p. 10-54.

Fig. 118-6. Harris, Analog Product Data Book, 1988, p. 10-96.

Fig. 118-7. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.

Chapter 119

Fig. 119-1. Electronics Digest, Spring 1988, p. 63.

Fig. 119-2. Electronic Engineering, 5/84, p. 36.

Fig. 119-3. Motorola, Linear Integrated, Circuits, p. 5-37.

Fig. 119-4. Reprinted from EDN, 8/18/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 119-5. Signetics, Linear Data Manual Vol. 3: Video, p. 10-58.

Fig. 119-6. Radio-Electronics, 2/86, p. 51.

Fig. 119-7. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-67.

Fig. 119-8. Radio-Electronics, 9/87, p. 41.

Fig. 119-9. Hands-On Electronics, Winter 1985, p. 44.

Fig. 119-10. Electronic Engineering, 8/86, p. 32.

Fig. 119-11. Signetics Linear Data Manual Vol. 3: Video, p. 11-95.

Fig. 119-12. Electronic Engineering, 7/84, p. 27.

Fig. 119-13. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-66.

Fig. 119-14. Reprinted from EDN, 5/16/85, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 119-15. Hands-On Electronics, 7-8/86, p. 78.

Fig. 119-16. Siliconix, Small-Signal FET Data Book, 1/86, p. 7-101.

Chapter 120

Fig. 120-1. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.

Fig. 120-2. Reprinted from EDN, 1/7/88, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 120-3. Hands-On Electronics, 10/88, p. 30.

Fig. 120-4. Reprinted with permission from Electronic Design. Copyright 1988, Penton Publishing.

Fig. 120-5. Gernsback Publications Inc., 42 New Ideas, 1984, p. 16.

Chapter 121

Fig. 121-1. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 47.

Fig. 121-2. Intersil, Component Data Catalog, 1987, p. 6-29.

Fig. 121-3. Reprinted from EDN, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 121-4. Intersil, Applications Handbook, 1988, p. 6-6.

Fig. 121-5. Reprinted from EDN, 10/1/87, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.

Fig. 121-6. Texas Instruments, Linear and Interface Circuit Applications, 1985, p. 7-19.

Fig. 121-7. Texas Instruments, Linear Circuits Data Book, 1989, p. 2-73.

Fig. 121-8. Popular Electronics, 10/89, p. 105.

Chapter 122

Fig. 122-1. Electronic Engineering, 11/76, p. 23.

Fig. 122-2. Datel, Data Conversion Components, p. 6-18.

Fig. 122-3. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 41.

Fig. 122-4. Motorola, Linear Integrated Circuits, p. 5-145.

Fig. 122-5. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN8-9.

Fig. 122-6. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 39.

Fig. 122-7. Maxim, 1986 Power Supply circuits, p. 44.

Fig. 122-8. Radio-Electronics, 4/85, p. 80.

Fig. 122-9. Motorola, Motorola TMOS Power FET Design Ideas, 1985, p. 36.

Chapter 123

Fig. 123-1. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN9-14.

Fig. 123-2. Analog Devices, Data Acquisition Databook, 1982, p. 6-27.

Fig. 123-3. Electronic Engineering, 12/75, p. 11.

Fig. 123-4. Electronic Engineering, 7/76, p. 23.

Fig. 123-5. Reprinted with permission from Electronic Design. Copyright 1975, Penton Publishing.

Fig. 123-6. Linear Technology Corp., Linear Databook, 1986, p. 5-79.
Fig. 123-7. Reprinted with permission from Electronic Design. Copyright 1975, Penton Publishing.
Fig. 123-8. Linear Technology Corp., Linear Databook, 1986, p. 5-33.
Fig. 123-9. Electronic Engineering, 10/77, p. 17.
Fig. 123-10. Linear Technology Corp., Linear Applications Handbook, 1987, p. AN9-13.
Fig. 123-11. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 2/87, p. 7-63.
Fig. 123-12. Reprinted with permission from Electronic Design. Copyright 1967, Penton Publishing.

Chapter 124

Fig. 124-1. Electronic Engineering, 10/84, p. 41.
Fig. 124-2. Siliconix, Integrated Circuits Data Book, 3/85, p. 3-21.
Fig. 124-3. National Semiconductor Corp., CMOS Databook, 1981, p. 3-41.
Fig. 124-4. Reprinted with permission from Electronic Design. Copyright 1970, Penton Publishing.
Fig. 124-5. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 16.
Fig. 124-6. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 16.
Fig. 124-7. Radio-Electronics, 11/82, p. 92.
Fig. 124-8. Teledyne, Teledyne Semiconductor Databook, p. 9.
Fig. 124-9. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 15.
Fig. 124-10. TAB Books, Third Book of Electronic Projects, p. 37.
Fig. 124-11. Ham Radio, 7/89, p. 62.
Fig. 124-12. Hands-On Electronics, 4/87, p. 93.
Fig. 124-13. Popular Electronics/Hands-On Electronics, 4/89, p. 25.
Fig. 124-14. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 11.
Fig. 124-15. Gernsback Publications Inc., 44 New Ideas, 1985, p. 8.
Fig. 124-16. Electronics Today International, 10/78, p. 95.
Fig. 124-17. Radio-Electronics, 1/86, p. 104.
Fig. 124-18. National Semiconductor Corp., Transistor Databook, 1982, p. 7-26.
Fig. 124-19. Electronic Engineering, 9/85, p. 25.
Fig. 124-20. Electronics Today International, 6/76, p. 42.
Fig. 124-21. National Semiconductor Corp., Linear Databook, 1982, p. 362.
Fig. 124-22. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 11/14/86, p. 5-269.

Chapter 125

Fig. 125-1. Reprinted with permission from Electronic Design. Copyright 1989, Penton Publishing.
Fig. 125-2. Popular Electronics, 6/89, p. 22.
Fig. 125-3. Electronic Engineering, Applied Ideas, 3/89, p. 32.

Chapter 126

Fig. 126-1. Electronic Engineering, 7/88, p. 27.
Fig. 126-2. Electronics Today International, 6/76, p. 40.
Fig. 126-3. Signetics, Analog Data Manual, 1982, p. 8-10.
Fig. 126-4. Reprinted from EDN, 5/73, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 126-5. Electronic Engineering, 1/83, p. 31. .
Fig. 126-6. Reprinted from EDN, 8/20/78, © 1989 Cahners Publishing Co., a division of Reed Publishing USA.
Fig. 126-7. Fairchild Camera and Instrument Corp., Linear Databook, 1982, p. 4-180.

Index

Numbers preceded by a "I," "II," "III," or "IV" are from *Encyclopedia of Electronic Circuits* Vol. I, II, III, or IV, respectively.

B

C

D

G

J

K

L

M

Q

R

S

T

U

V

W

X

Y

Z

Rudolf F. Graf & William Sheets

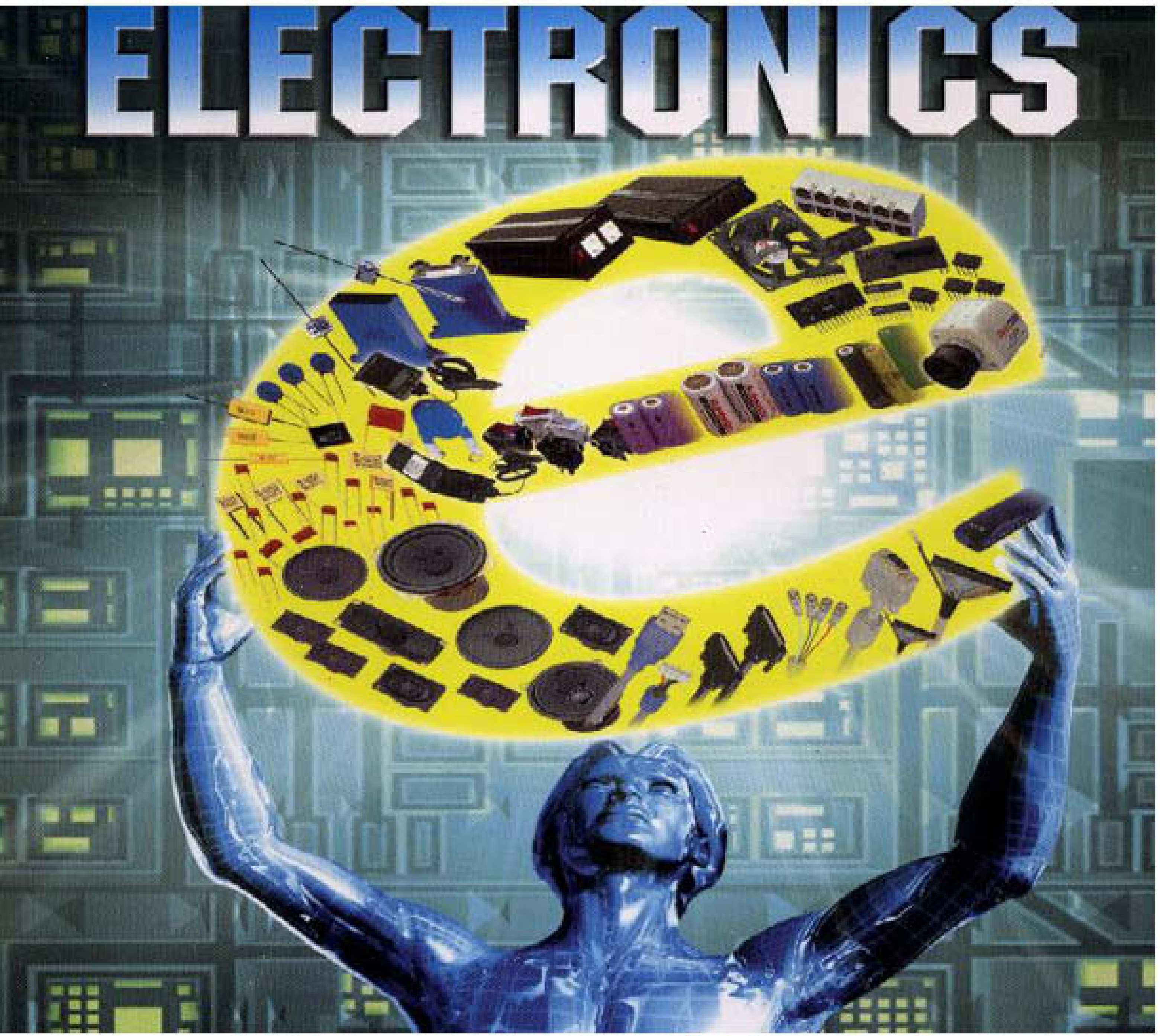
Encyclopedia of Electronic Circuits

VOLUME 4

Encyclopedia of Electronic Circuits

VOLUME 4

Patent notice


ELECTRONICS

Encyclopedia of Electronic Circuits

VOLUME 4

Rudolf F. Graf & William Sheets

TAB

TAB BOOKS

Blue Ridge Summit, PA

FIRST EDITION
FIRST PRINTING

Published by TAB Books.
TAB Books is a division of McGraw-Hill, Inc.

Library of Congress Cataloging-in-Publication Data

(Revised for volume 4)

Graf, Rudolf F.
Encyclopedia of electronic circuits.

Vol. 4: By Rudolf F. Graf and William Sheets.
Includes bibliographical references and indexes.
1. Electronic circuits—Encyclopedias. I. Sheets, William. II. Title.
TK7867.G66 1985 621.3815'.3 84-26772
ISBN 0-8306-0938-5 (v. 1)
ISBN 0-8306-1938-0 (pbk. : v. 1)
ISBN 0-8306-3138-0 (pbk. : v. 2)
ISBN 0-8306-3138-0 (v. 2)
ISBN 0-8306-3348-0 (pbk. : v. 3)
ISBN 0-8306-7348-2 (v. 3)
ISBN 0-8306-3895-4 (pbk. : v. 4)
ISBN 0-8306-3896-2 (v. 4)

Acquisitions Editor: Roland S. Phelps
Technical Editor: Andrew Yoder
Director of Production: Katherine G. Brown
Cover Design: Holberg Design, York, PA

TAB1

Contents

Introduction

Volume 4 of *Encyclopedia of Electronic Circuits* contains many new, not previously covered circuits, organized into 104 chapters. Circuit titles are listed at the beginning of each chapter, for references. Most of these circuits appeared in publications since 1988 and should be very useful for obtaining new ideas for research and development, or simply to fill a need for a specific circuit idea or application. Those wishing to develop their own circuits will find this book indispensable as a source of ideas, to see how others have solved a problem or approach a design, and to obtain a starting point toward a new design.

A brief explanation accompanies almost every entry. Those that have been omitted are either repetitive, obvious, or too involved to describe in few words. In this case, the reader should consult the original sources (as listed in the back of the book).

We also wish to extend our sincere thanks to Ms. Loretta Gonsalves for her fine work at the word processor. Her skill and cooperation contributed much to the successful completion of this book.

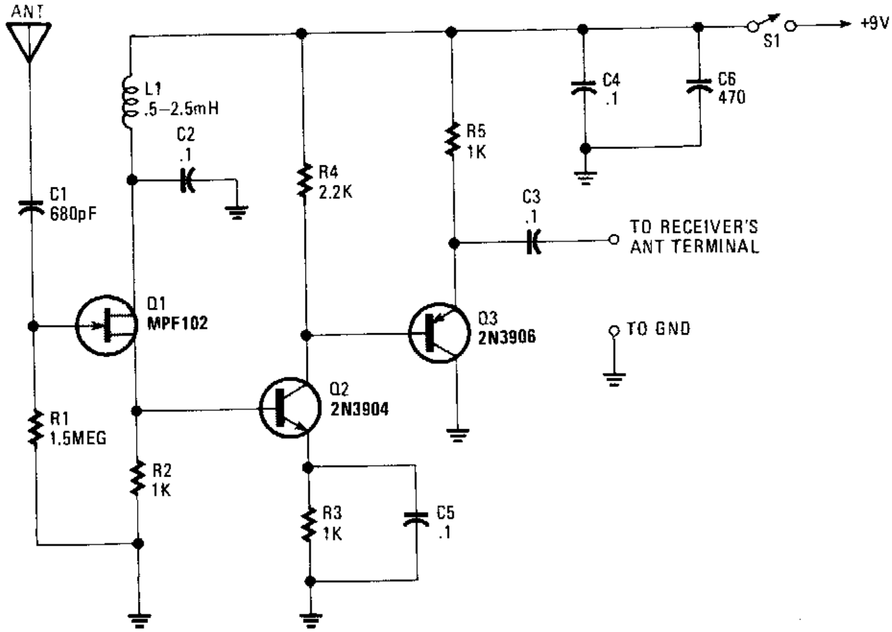
Rudolf Graf and William Sheets

1

Active Antennas

The sources of the following circuits are contained in the Sources section, which begins on page 660. The figure number in the box of each circuit correlates to the entry in the Sources section.

ACTIVE ANTENNA WITH GAIN



POPULAR ELECTRONICS

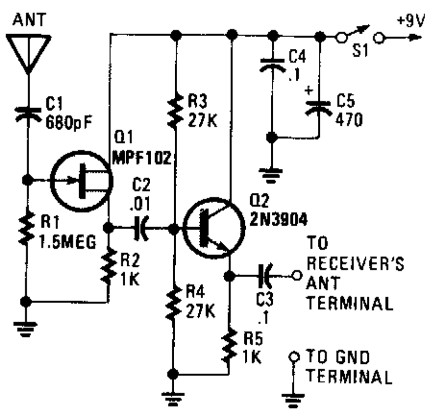
Fig. 1-1

The signal booster, built around a few transistors and support components, offers an RF gain of about 12 to 18 dB (from about 100 kHz to over 30 MHz).

The RF signal is direct-coupled from Q1's source terminal to the base of Q2, which is configured as a voltage amplifier. The output of Q2 is then direct-coupled to the base of Q3 (configured as an emitter-follower amplifier). Transistor Q3 is used to match and isolate the gain stage from the receiver's RF-input circuitry.

Inductor L1 is used to keep any power source noise from reaching the FET (Q1) and any value of RF choke from 0.5 to 2.5 mH will do. The value of R2 sets the Q2 bias at about 2 V. If the voltage is less than 2 V, increase the value of R2 to 1.5 kΩ. To go below 100 kHz (to the bottom of the RF spectrum), increase the value of C1 to 0.002 μF. The antenna is a short pull-up type (42″ to 86″ long).

ACTIVE ANTENNA I

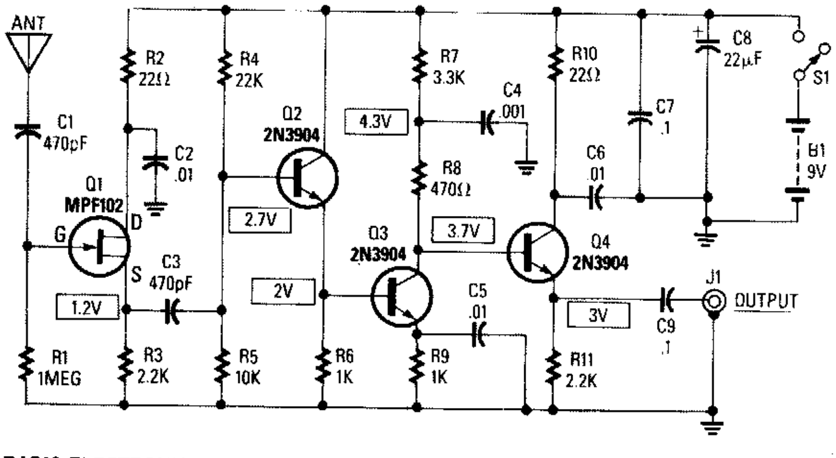


POPULAR ELECTRONICS *Fig. 1-2*

This circuit is designed to make a short pull-up antenna perform like a long wire antenna, while offering no voltage gain. The circuit boosts the receiver's performance only if the signal at the antenna is of sufficient level to begin with.

This circuit takes a short pull-up antenna that has a high output impedance and couples it to the receiver's low input impedance through a two-transistor impedance-matching network. Transistor Q1's high input impedance and high-frequency characteristics make it a good match for the short antenna, and Q2's low output impedance is a close match for the receiver's input. This circuit is usable over the range from 100 kHz to 30 MHz.

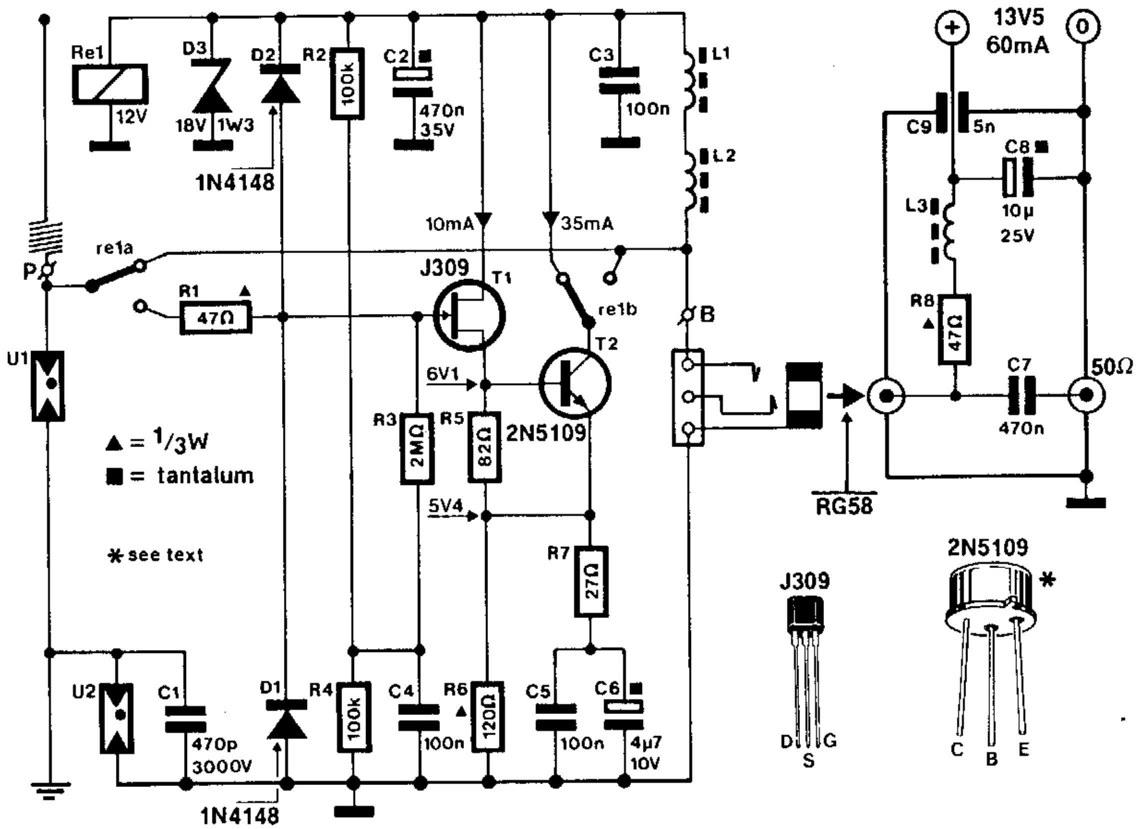
ACTIVE ANTENNA II



RADIO-ELECTRONICS *Fig. 1-3*

This circuit provides 14- to 20-dB gain at frequencies from 10 kHz to 30 MHz. The antenna length can be anything between 5 and 10 feet. A 102-inch CB whip is excellent for this purpose.

WIDEBAND ACTIVE ROD ANTENNA



ELEKTOR ELECTRONICS

Fig. 1-4

A J309 Siliconix FET feeds a 2N5109 in a wideband RF amplifier configuration. A relay is used to bypass the amplifier in the transmit mode (if desired). A 2-m 5/8-wave whip is used as the active antenna element. The amplifier is fed dc via the coax cable, which makes the use of only a single coax lead for both signal and power. U1 is a surge arrester for electrostatic discharge protection.

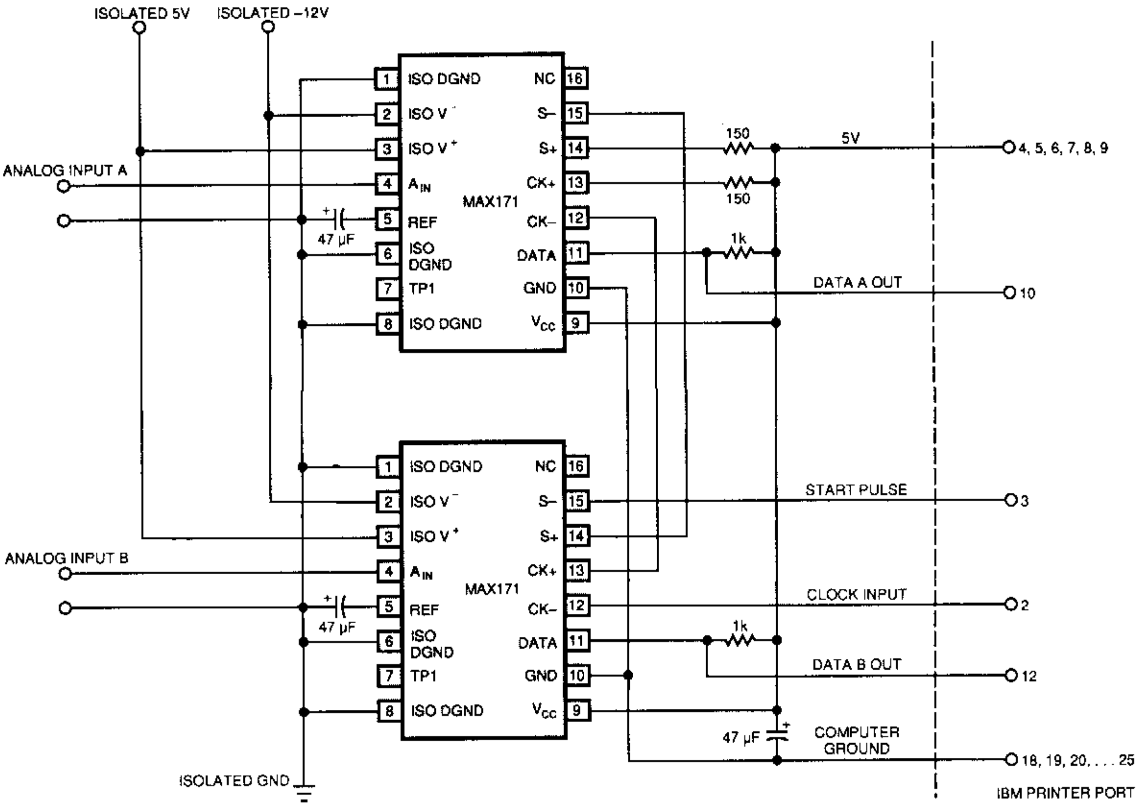
2

Analog-to-Digital Converter

The source of the following circuit is contained in the Sources section, which begins on page 660. The figure number in the box correlates to the entry in the Sources section.

A/D Board

A/D BOARD



EDN

Fig. 2-1

An IBM PC can operate the two 12-bit A/D converters in Fig. 1 via its printer port. The converters' serial outputs use only two of the printer port's eight data lines (DATA A OUT, DATA B OUT). Because the IBM PC's printer port supplies no power, interface software running on the PC programs the six unused data lines high. Busing these data lines provides power for the digital portion of the A/D converters. (The converters have internal optoisolators. Consequently, you must provide isolated supplies for their analog sides.)

Although the converters can execute 12-bit conversions in 6 μs, the slow software-driven approach used in this Design Idea stretches conversion periods out to about 100 μs (depending on your PC's clock speed).

The circuit takes advantage of the converters' optoisolator inputs to put their clock and start inputs in series. Therefore, the converters operate synchronously.

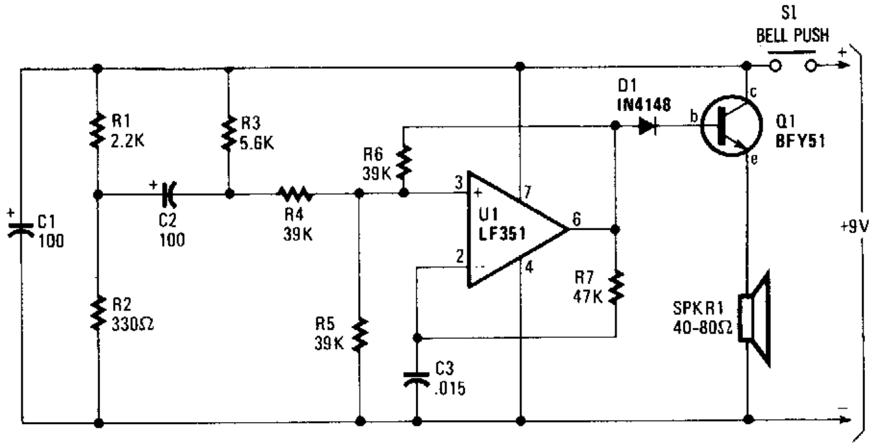
The accompanying software starts the conversions, issues clock pulses, reads the data bits as they become available, and stores them in memory. The listing is too long to reproduce here; you can obtain it from the EDN BBS (617−558−4241, 2400, 8, N, 1).

3

Annunciators

The sources of the following circuits are contained in the Sources section, which begins on page 660. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC DOOR BUZZER

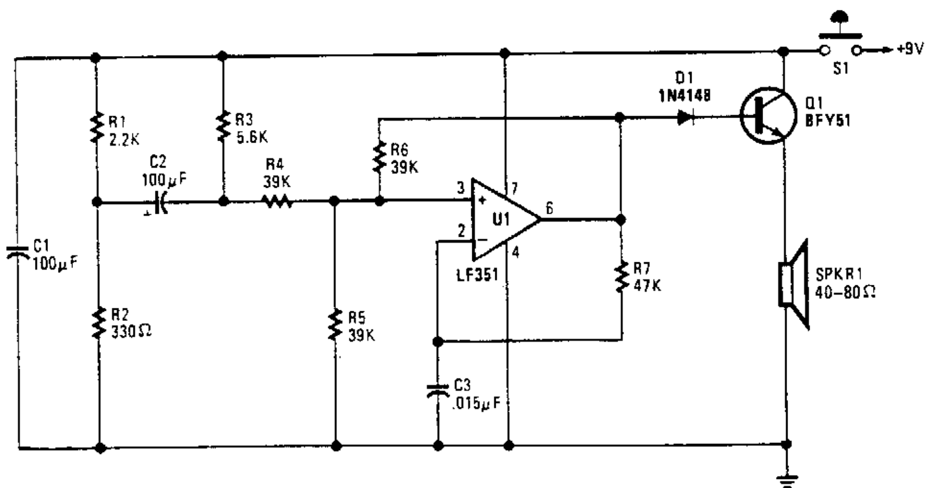


HANDS-ON ELECTRONICS

Fig. 3-1

When S1 is depressed, an initial positive voltage is placed on C2 and the noninverting terminal of U1. The circuit oscillates at a low frequency. As C2 charges up through R3, a rapid increase in frequency of oscillation results, producing (at SPKR1) a rapidly rising pitched sound. This sound is easily recognized over ambient noise.

DOOR BUZZER

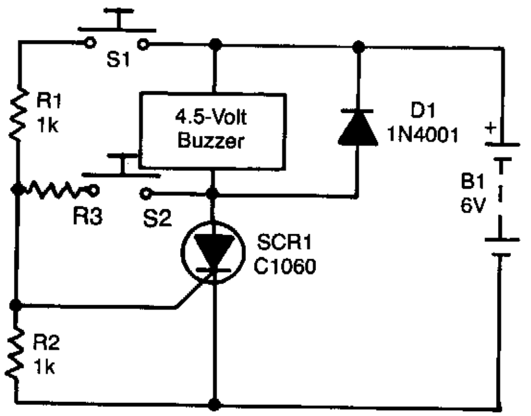


POPULAR ELECTRONICS

Fig. 3-2

An LF357 functions as a swept-tone oscillator, driving Q1 and SPKR1. A 9-Vdc supply is required.

SCR CIRCUIT WITH SELF-INTERRUPTING LOAD



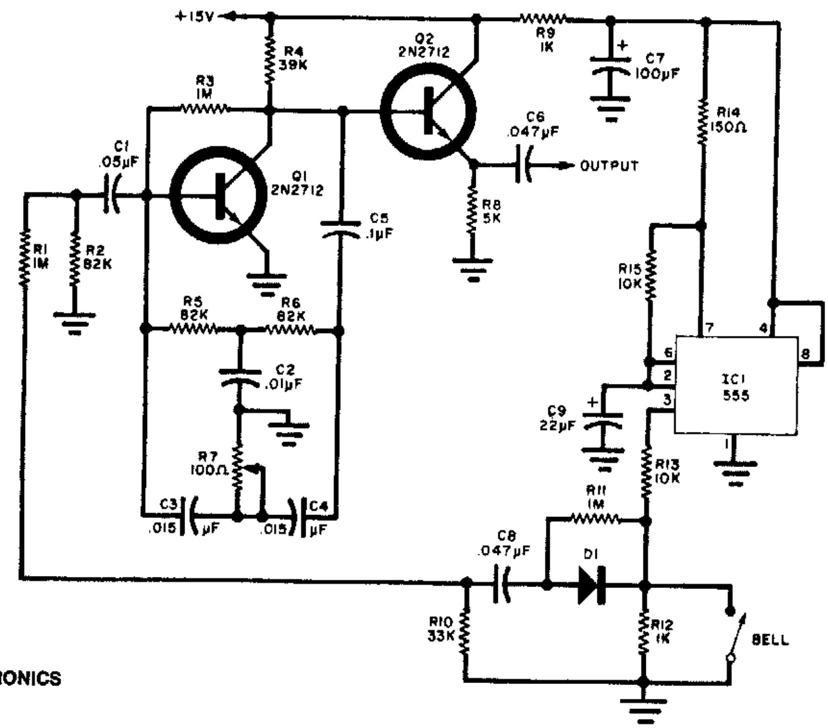
RADIO-ELECTRONICS

Fig. 3-3

A self-interrupting device connected to a voltage source functions as a switch that repeatedly opens and closes; therefore, the circuit does not latch in the normal way, so the alarm operates only as long as S1 is closed. Because of the inductive nature of that type of load, a damping diode (D1) must be wired across it.

The circuit can be modified to provide a self-latching action simply by wiring a 470-Ω resistor in parallel with the alarm. The circuit latches because the anode current of the SCR does not fall to zero when the alarm self-interrupts, but to a value that is determined by the value of the R3. The circuit can be unlatched by pressing S2, thereby enabling the anode current to fall to zero when the alarm self-interrupts.

ELECTRONIC BELL

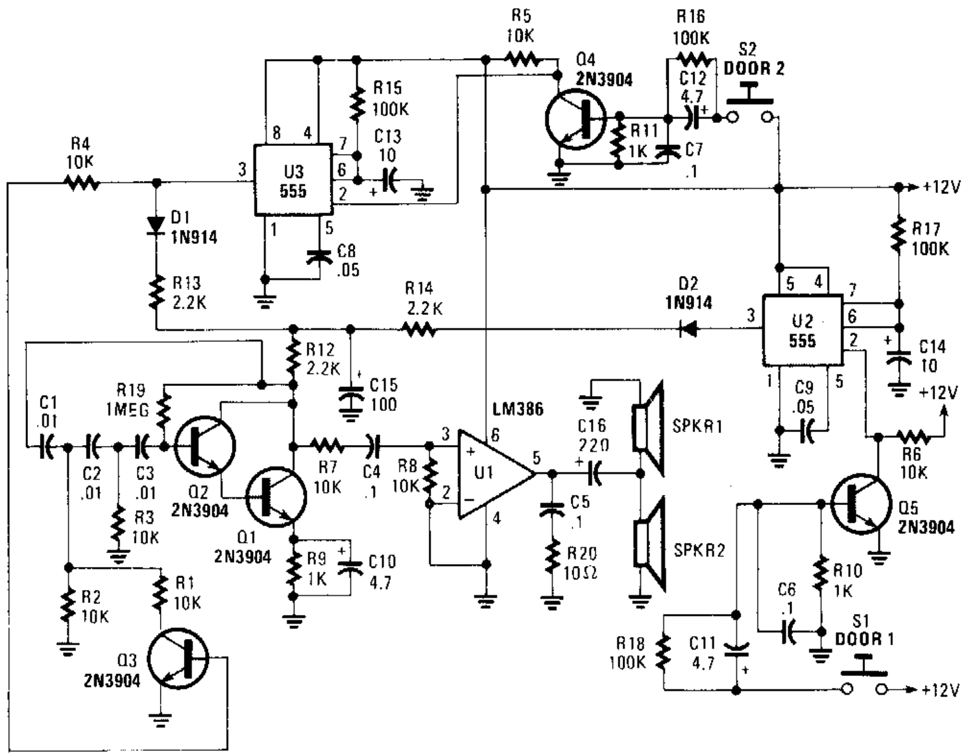


POPULAR ELECTRONICS

Fig. 3-4

A 555 timer pulses twin-T oscillator Q1. Q2 acts as an output buffer. R7 adjusts the frequency of oscillator Q1.

TWO-DOOR ANNUNCIATOR



POPULAR ELECTRONICS ***Fig. 3-4***

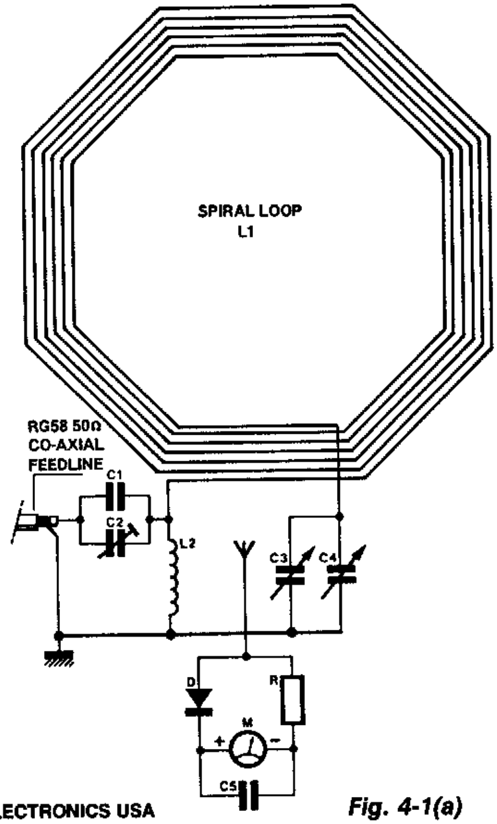
When the pushbuttons at either door are depressed, this circuit generates a different tone for each door. Tones are generated by phase-shift oscillator Q1/Q2. Q3 provides tone frequency change by changing the phase-shift network. U2 and U3 are timers for the tones and Q4/Q5 interface the timers with the pushbuttons.

4

Antenna Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 660. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOOP ANTENNA FOR 3.5 MHz

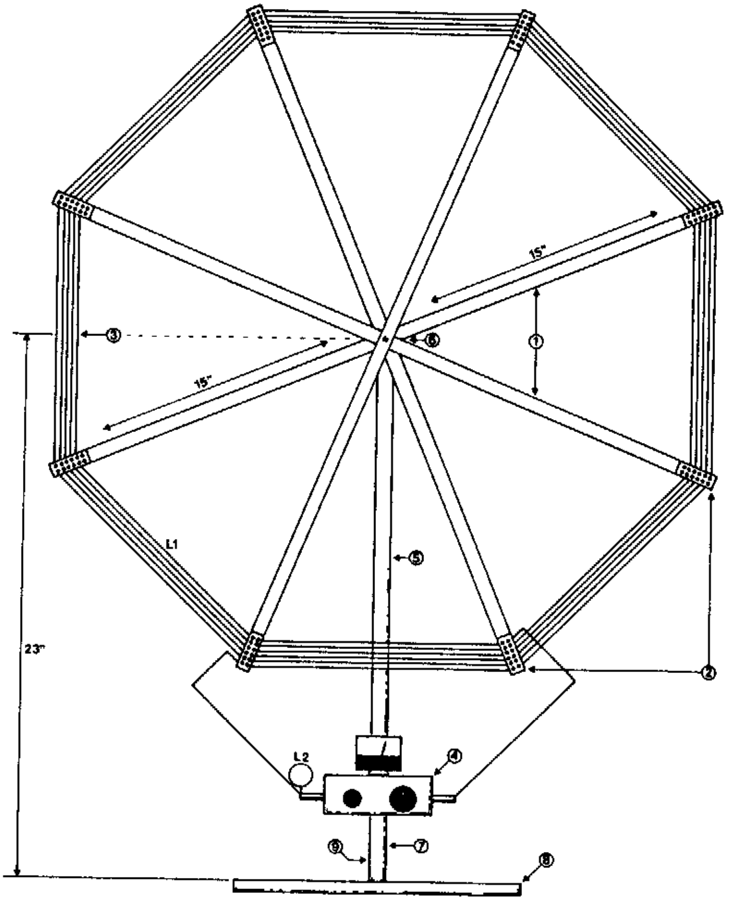


ELEKTOR ELECTRONICS USA

Fig. 4-1(a)

COMPONENTS LIST

C1 = 3 750 pF 500 V silver-mica capacitor.
C2 = 100 pF preset capacitor (Jackson C803).
C3 = 75 pF variable capacitor (Jackson C809), plus knob.
C4 = 12.7 pF variable capacitor (Jackson C16), plus knob.
C5 = 22 nF mica capacitor.
M = 250 μA f.s.d. 40 x 40 mm moving coil meter (Maplin LB808).
D = HF silicon diode.
R = 1 kΩ resistor (see text).
L1 = 5 1/8 turns of PVC covered stranded 7/0.2 mm wire. Outside diameter: 1.2 mm, 1 kV/1.5 A rating (see text).
L2 = 13 turns 16SWG tinned wire, 1 inch internal diameter.
Feedline = 48 inch RG58 coaxial cable, plus plug to suit transmitter.
Box = ABS box type MB3, 118 x 96 x 45 mm. Maplin ref. LH22.
Terminal blocks = qty. 4 12-way 2 amp terminal block. Maplin ref. FE78.
Spacers = qty. 3 insulated spacer type M3, 30 mm long, Maplin ref. FS40T.
Spokes = qty. 4 8-foot lengths of 5/8 x 1/4 inch molded hardwood (DIY store).
Vertical support = 23 x 0.8 x 0.8 inch wood (DIY store).
Wood base = 12 x 8 x 0.5 inch plywood or similar.
2 1/2 inch steel support bracket.



ELEKTOR ELECTRONICS USA

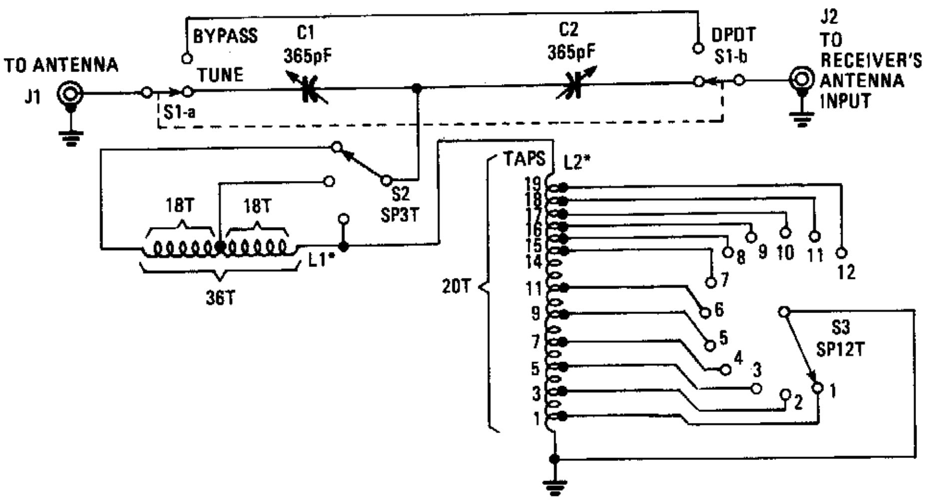
Fig. 4-1(b)

1. 4 lengths molded hardwood 30″ × 5/8 × 1/4″. Varnished. 2BA holes drilled in the centre. Glued and bolted together.
2. 8 off 6-way 2-amp polythene terminal blocks used as insulated wire spacers.
3. 51/8 turns of PVC stranded wire (for specs see components list).
4. See Fig. 3.
5. Wood vertical support 23″ × 0.8″ × 0.8″, wood stained.
6. 2″ × 2BA bolt.
7. Box front vertical support, 41/2″ × 1/2″ × 3/4″, wood stained.
8. Wood base 12″ × 8″ × 1/2″ (for similar), wood stained.
9. 21/2 steel support bracket behind wood vertical support.
10. Drilled and secured with glue and c/s wood screws.

Note: ″ = inch = 2.54 cm.

Suitable for receiving or transmitting (10 W or less) on the 80-m band, this loop antenna might be helpful when an outside antenna is not possible.

1-TO 30-MHz ANTENNA TUNER



POPULAR ELECTRONICS

Fig. 4-2

L1 = 36T #18 enamel wire
on 2″ PVC SCH 40 pipe.

L2 = 20T #18 enamel (as in L1)
tapped as shown C1/C2.

365-pF variable capacitor,
receiving type.

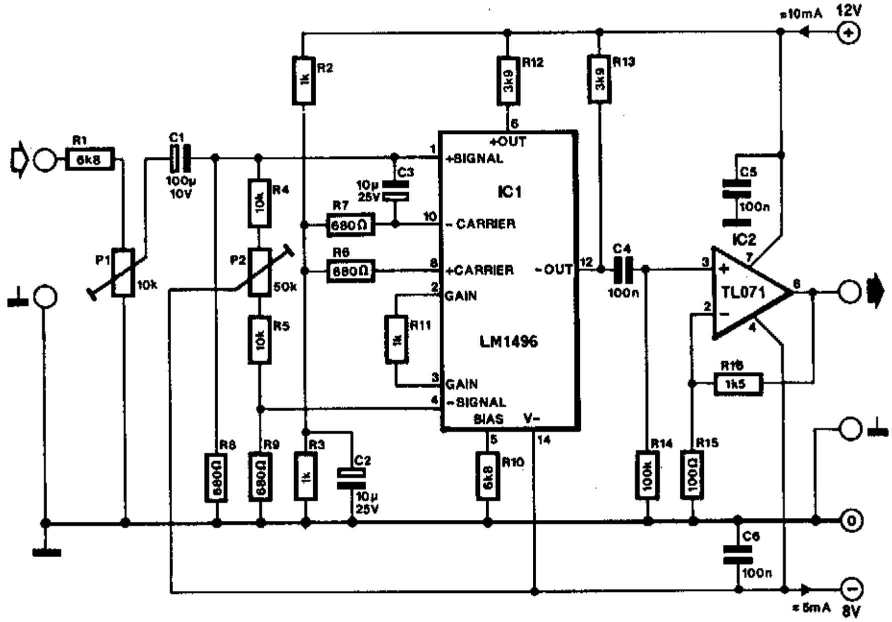
This tuner will match a random length wire antenna to a receiver or low-power transmitter (≤25 W) for optimum signal transfer.

5

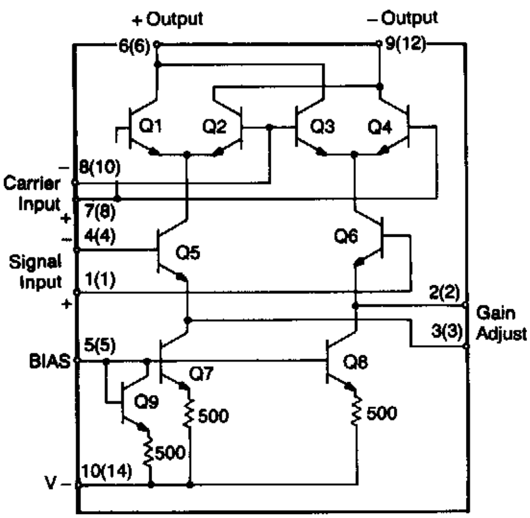
Audio Effects Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 660. The figure number in the box of each circuit correlates to the entry in the Sources section.

AUDIO-FREQUENCY DOUBLER



HANDS-ON ELECTRONICS



Internal circuit of the LM1496.

ELEKTOR ELECTRONICS

Fig. 5-1

Often the frequency of a signal must be doubled: modulator/demodulator chip LM1496 is an ideal basis for this.

From trigonometry it is well known that:

$$2\sin x\cos x = \sin 2x$$

and:

$$\sin^2 = 1 - x\cos 2x.$$

These equations indicate that the product of two pure sinusoidal signals of the same frequency is one signal of double that frequency. The purity of the original signals is important: composite signals would give rise to all sorts of undesired products.

The LM1496 can only process signals that are not greater than 25 mV: above that level, serious distortion will occur. The design is therefore provided with a potential divider at its input. This addition makes it possible, for instance, to arrange for a 500-mV input signal to result in a signal of only 25 mV at the input of the LM1496.

To provide a sufficiently high output signal, the output of IC1 is magnified by op amp IC2, which is connected as a noninverting amplifier. Because the output of IC1 contains a dc component of about 8 V, the coupling between the two stages must be via a capacitor, C4.

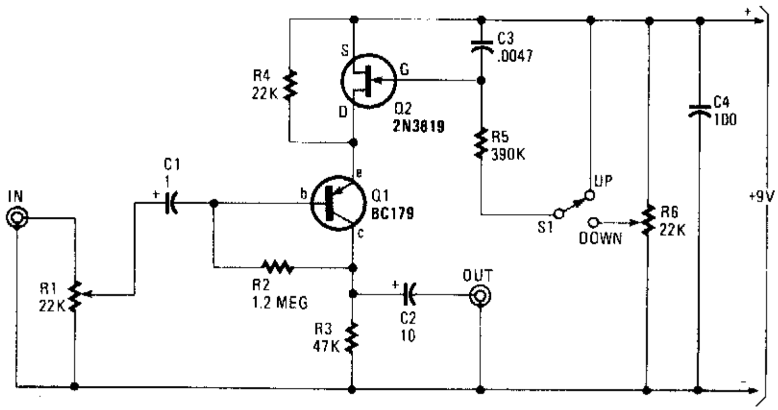
With values of R15 and R16 as shown, IC2 gives an amplification of 16 (24 dB). The overall amplification of the circuit depends on the level of the input signal: with an input of 1.2 V, the amplification is unity; when the input drops to 0.1 V, the amplification is just 0.1. The value of the input resistors has been fixed at 680 Ω: this value gives a reasonable compromise between the requirements for a high input impedance and a low noise level.

To ensure good suppression of the input signal at the output, the voltages at pin 1 and pin 4 of IC1 must be absolutely identical to P4. It is possible, with the aid of a spectrum analyzer, to suppress the fundamental (input) frequency by 60 to 70 dB.

The output signal at pin 12 is distorted easily, because the IC is not really designed for this kind of operation. The distortion depends on the level of the input signal. At a frequency of 1 kHz and an input level of 100 mV, the distortion is about 0.6%; when the input level is raised to 500 mV, the distortion increases to 2.3%, and when the input level is 1 V, the distortion is 6%. The signal-to-noise ratio under these conditions varies between 60 and 80 dB.

The circuit draws a current of 10 mA from the positive supply line and 5 mA from the negative line. The phase shift between the input and output signals is about 45° (output lags). Finally, although the normal output is taken from pin 12, a similar output that is shifted by 180° (with respect to that at pin 12), is available at pin 6.

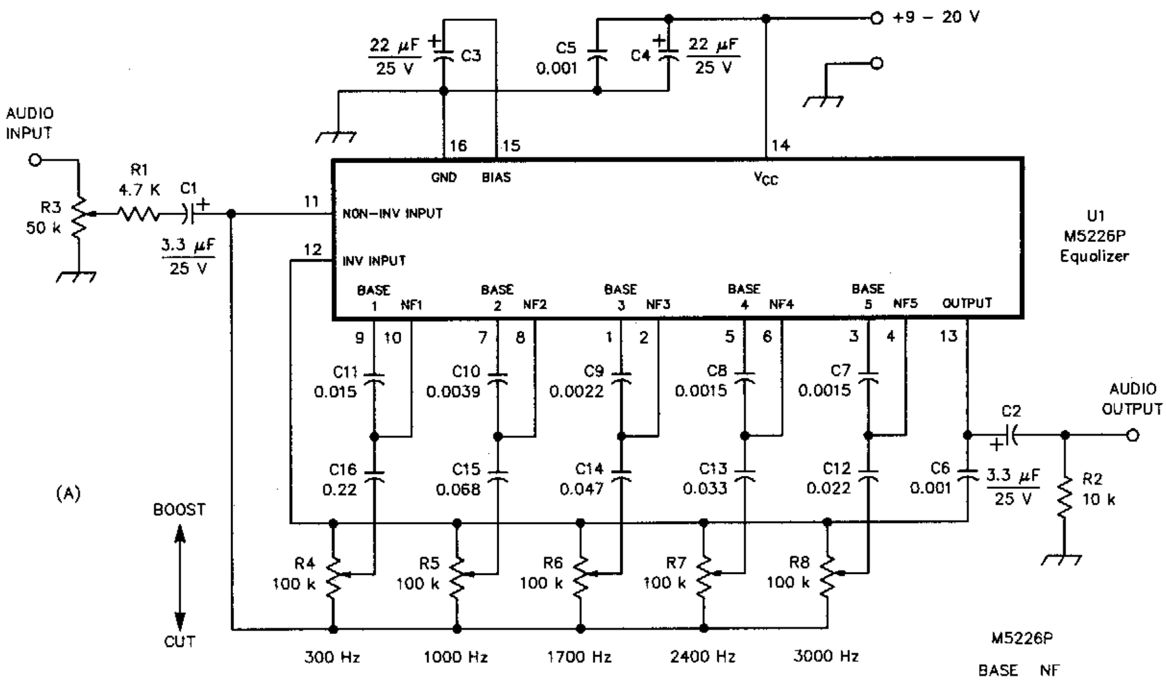
AUDIO FADER

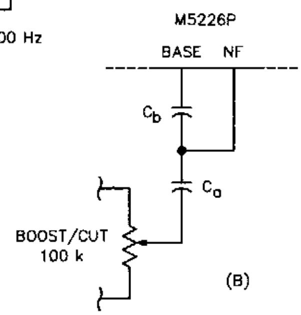


HANDS-ON ELECTRONICS *Fig. 5-2*

In this circuit, Q1 is a simple amplifier that has its gain controlled by a variable emitter resistance supplied by FET Q2. In the up position of S1, C3 discharges through R5 and the gain of Q1 decreases because Q2 is driven toward cut-off. In the down position, Q2 conducts more, depending on the setting of R6, which causes a gain increase. By varying R5 or C3, various fade rates can be obtained.

AUDIO EQUALIZER





PC−Board Connections

Terminal	Connector Finger
AUDIO INPUT	5, 8, 14, 21
AUDIO OUTPUT	4, 7, 13, 20
+9 − 20 V	2
(ground)	1, 6, 9, 12, 11, 6, 19, 22

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000.

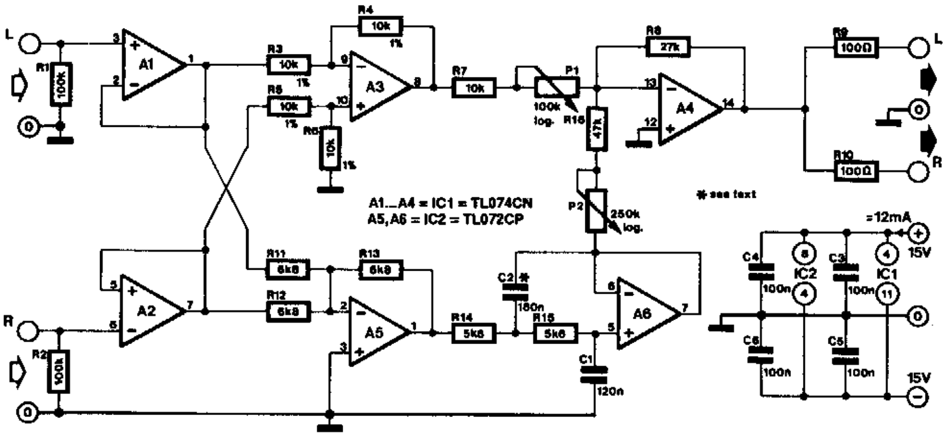
This schematic represents one of four sections (channels) on the PC board.

QST

Fig. 5-3

Designed for communications use, this equalizer circuit uses a Mitsubishi M5226P audio equalizer IC to adjust frequency response. It runs from a 9 to 20 V supply. C6 through C16 are polyester film capacitors of ± 5% tolerance.

VOCAL ELIMINATOR



ELEKTOR ELECTRONICS *Fig. 5-4*

Otherwise properly mixed sounds often suffer from a predominant solo voice (which might, of course, be the intention). If such a voice needs to be suppressed, the present circuit will do the job admirably.

The circuit is based on the fact that solo voices are invariably situated "at the center" of the stereo recordings that are to be mixed. Thus, voice levels in the left- and right-hand channels are about equal. Arithmetically, therefore, left minus right equals zero; that is, a mono signal without voice.

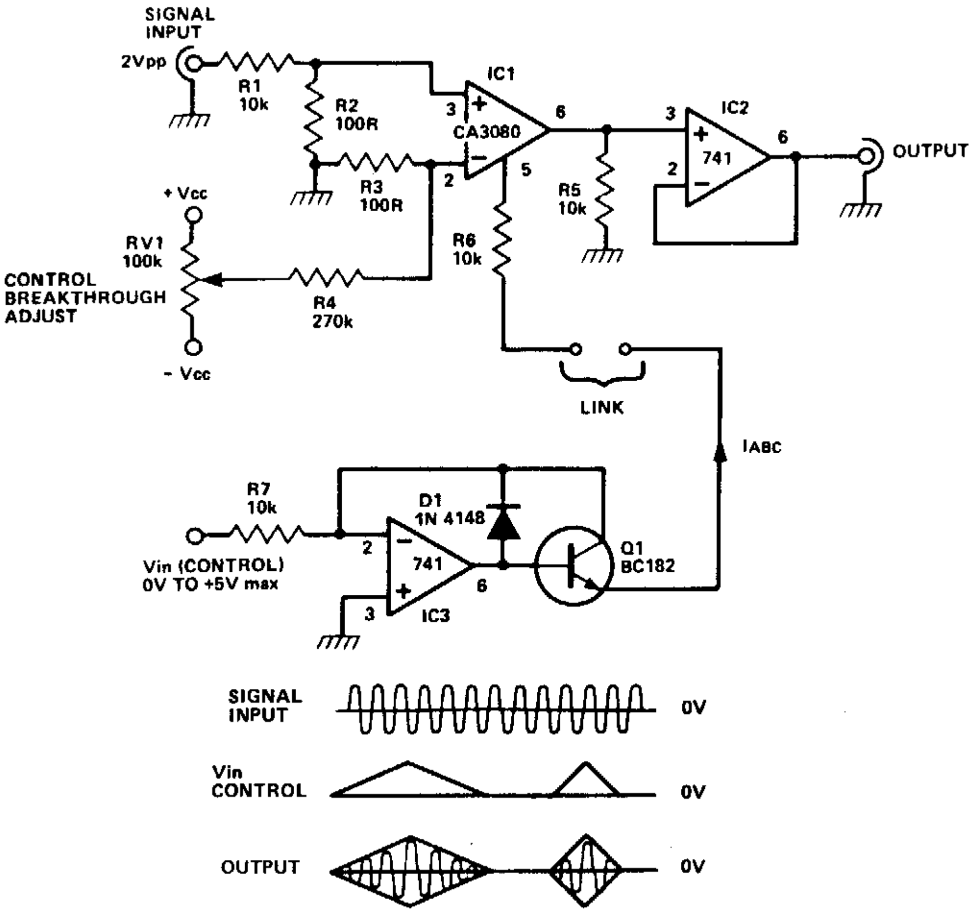
There is, however, a problem: the sound levels of bass instruments, more particularly the double basses, are also just about the same in the two channels. On the one hand low-frequency sounds are virtually nondirectional and on the other hand, the recording engineers purposely use these frequencies to give a balance between the two channels.

However, the bass instruments can be recovered by adding those appearing in the left + right signal to the left − right signal. The whole procedure is easily followed in the circuit diagram. The incoming stereo signal is buffered by A1 and A2. The buffered signal is then fed to differential amplifier A3 and subsequently to summing amplifier A5. The latter is followed by a low-pass filter formed by A6. You can choose between a first-order and a second-order filter by respectively omitting or fitting C2. Listen to what sounds best.

The low-frequency signal and the difference signal are applied to summing amplifier A4. The balance between the two is set by P1 and P2 to individual taste.

You have noticed that the circuit does not contain input or output capacitors. If you wish, output capacitors can be added without detriment. However, adding input capacitors is not advisable, because the consequent phase shift would adversely affect the circuit operation.

VOLTAGE-CONTROLLED AMPLIFIER



ELECTRONICS TODAY INTERNATIONAL

Fig. 5-5

The CA3080 can be used as a gain controlling device. The input signal is attenuated by R1/R2 so that a 20-mVpp signal is applied to the input terminals. If this voltage is much larger, significant distortion will occur at the output. In fact, this distortion is put to good use in the triangle-to-sine wave converter.

The gain of the circuit is controlled by the magnitude of the current IABC. This current flows into the CA3080 at pin 5, which is held at one diode voltage drop above the $-V_{CC}$ rail. The gain of the CA3080 is "linearly" proportional to the magnitude of the IABC current over a range of 0.1 μA to 1 mA. Thus, by controlling IABC, you can control the signal level at the output. The output is a current output, which has to be "dumped" into a resistive load (R5) to produce a voltage output. The output impedance at IC1 pin 6 is 10 kΩ (R5), but this is "unloaded" by the voltage follower (IC2) to produce a low output impedance.

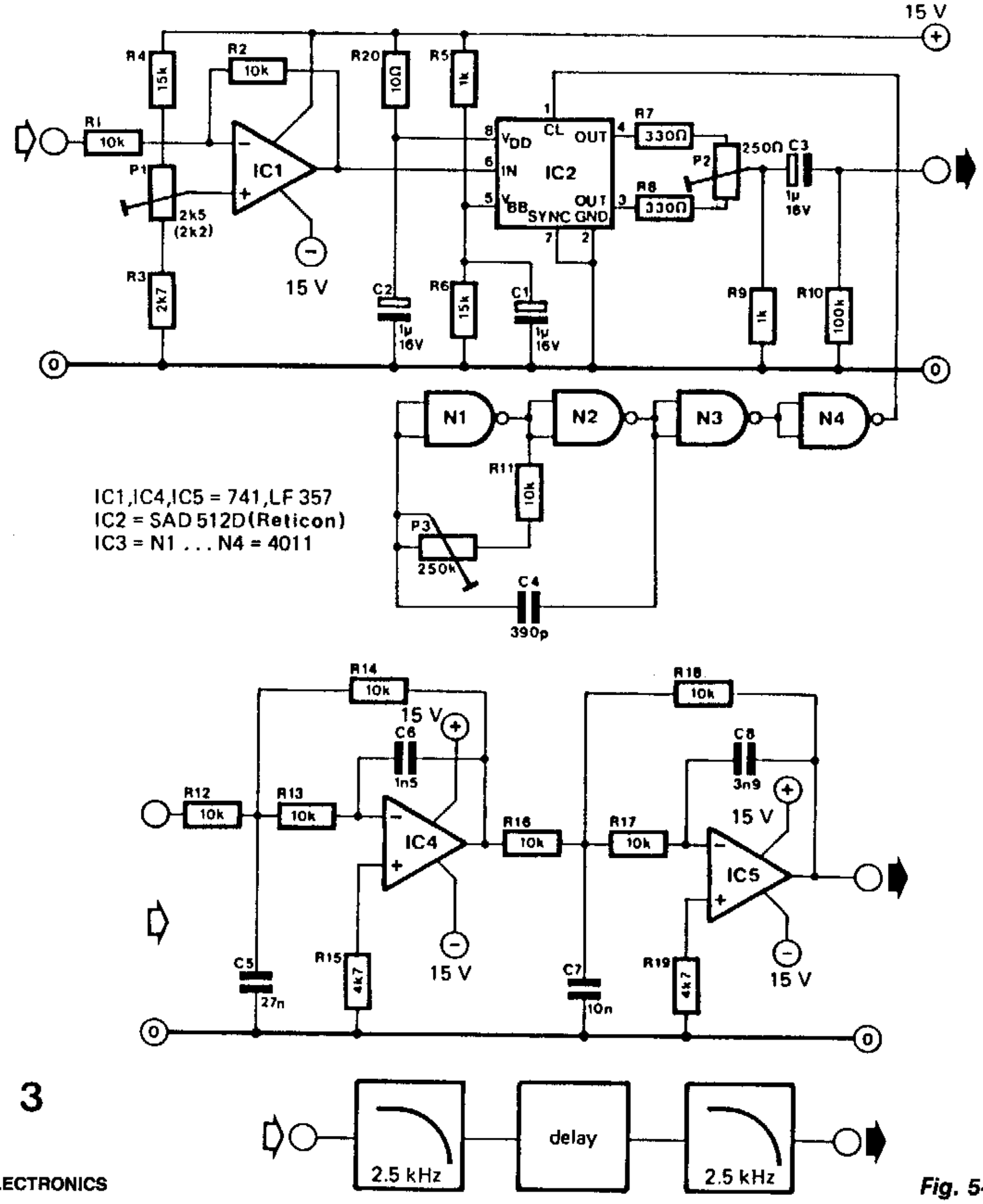
The circuit around IC3 is a precision voltage-to-current converter and this can be used to generate IABC. When V_{in} (control) is positive, it linearly controls the gain of the circuit. When it is negative, IABC is zero and so the gain is zero.

ANALOG DELAY LINE (ECHO AND REVERB)

1

2

3

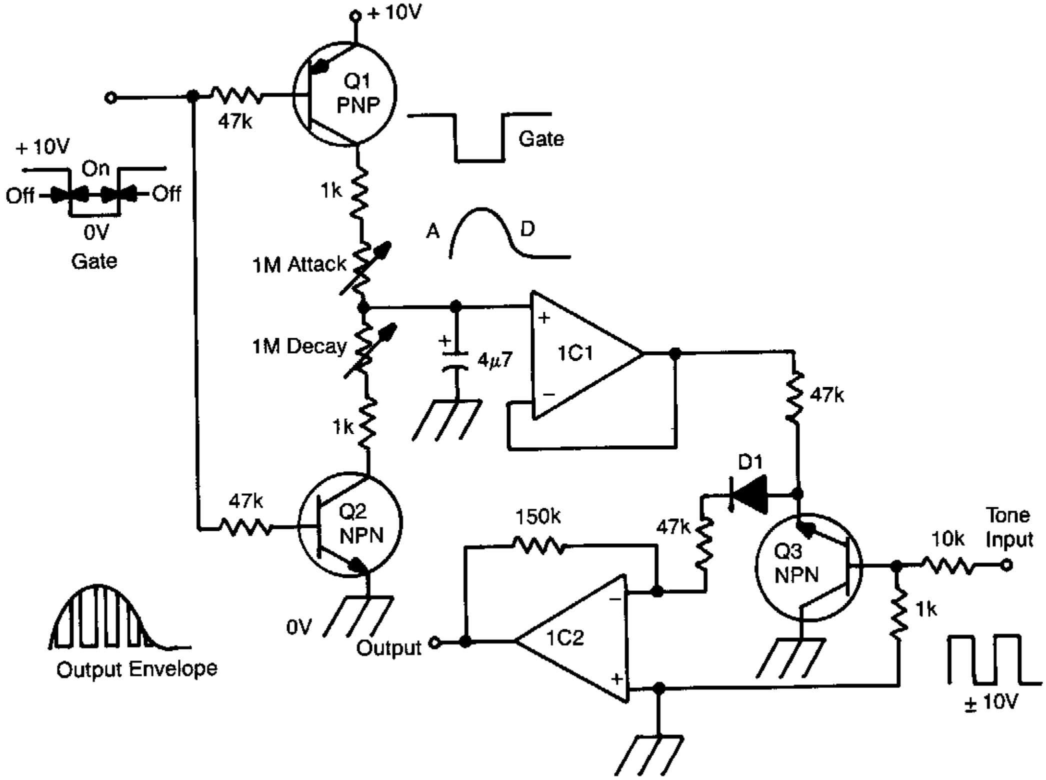


ELEKTOR ELECTRONICS

Fig. 5-6

This circuit uses an SAD 512D (Reticon) chip, which is a 512-stage analog shift register. By varying the clock frequency between 5 and 50 kHz, delay time can be set between 51.2 and 5.12 ms. The clock frequency must be at least twice the highest audio frequency.

MUSICAL ENVELOPE GENERATOR AND MODULATOR



ELECTRONICS TODAY INTERNATIONAL

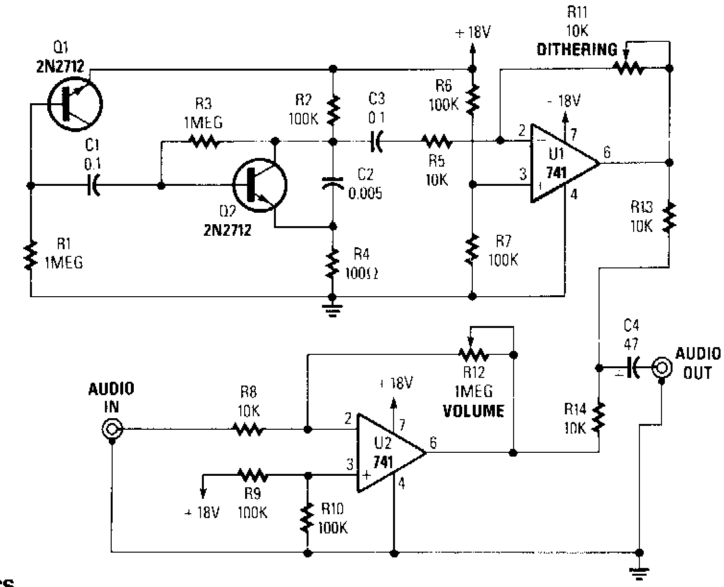
Fig. 5-7

A gate voltage is applied to initiate the proceedings. When the gate voltage is in the ON state, Q1 is turned on, and capacitor C is charged up via the attack pot in series with the 1-kΩ resistor. By varying this pot, the attack time constant can be manipulated. A fast attack gives a percussive sound, a slow attack gives the effect of "backward" sounds. When the gate voltage returns to its OFF state, Q2 is turned on and the capacitor is then discharged via the decay pot and the other 1-kΩ resistor to ground. Thus, the decay time constant of the envelope is also variable.

This envelope is buffered by IC1, a high-impedance voltage follower and is applied to Q3, which is being used as a transistor chopper. A musical tone in the form of a square wave is connected to the base of Q3. This turns the transistor on or off. Thus, the envelope is chopped up at regular intervals, which are determined by the pitch of the square wave.

The resultant waveform has the amplitude of the envelope and the harmonic structure of the square wave. IC2 is used as a virtual earth amplifier to buffer the signal and D1 ensures that the envelope dies away at the end of a note.

AUDIO DITHERIZING CIRCUIT FOR DIGITAL AUDIO USE



POPULAR ELECTRONICS

Fig. 5-8

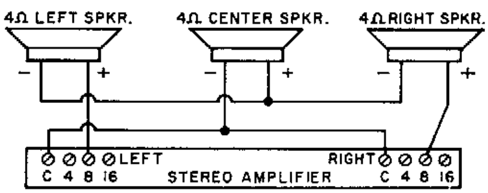
By adding a small amount of noise to a signal to be digitized (about 0.7 bit):

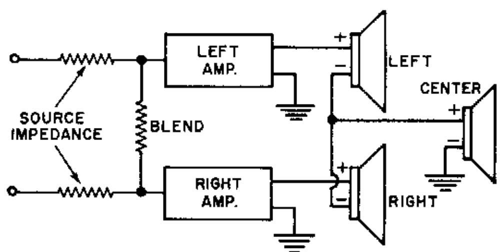
$$V_{\text{NOISE}} \approx 0.7 \left(\frac{V_{\text{INPUT}}\text{p-p}}{2^N} \right)$$

where: $n = \#$ of bits. For example, 8 bits and 2 V p-p would be 0.0055 V.

This circuit uses a transistor (Q1) and an amplifier (Q2 and U1) to generate the noise signal. R11 controls the noise injection and R12 controls the gain of the system.

DERIVED CENTER-CHANNEL STEREO SYSTEM





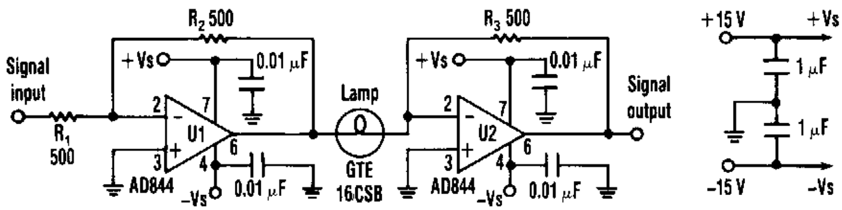
ELECTRONICS WORLD

Fig. 5-9(a)

Fig. 5-9(b)

A simple method of deriving a center or third channel without the use of an extra transformer or amplifier. (a) 4-Ω speakers are connected to 8-Ω amplifier taps. 8 and 16-Ω speakers connect to 16-Ω taps. (b) By blending the inputs it is possible to cancel out undesired crosstalk.

LOW-DISTORTION AMPLIFIER/COMPRESSOR



ELECTRONIC DESIGN *Fig. 5-10*

Designers can build a 15-dB compressor with a miniature lamp and a current-feedback amplifier. The circuit possesses extremely low distortion at frequencies above the lamp's thermal time constant. This means that distortion is negligible from audio frequencies to beyond 10 MHz. There's also relatively little change in phase versus gain compared to other automatic gain-control circuits. Lastly, the circuit has many instrumentation, audio, and high-frequency applications as a result of its low distortion and small phase change.

The AD844 op amp is a perfect fit for this application because it's a current-feedback amplifier. Each stage of the circuit, U2, lamp, and feedback resistor compresses an ac signal by over 15 dB (see the figure). Cascading a number of stages delivers higher compression ranges.

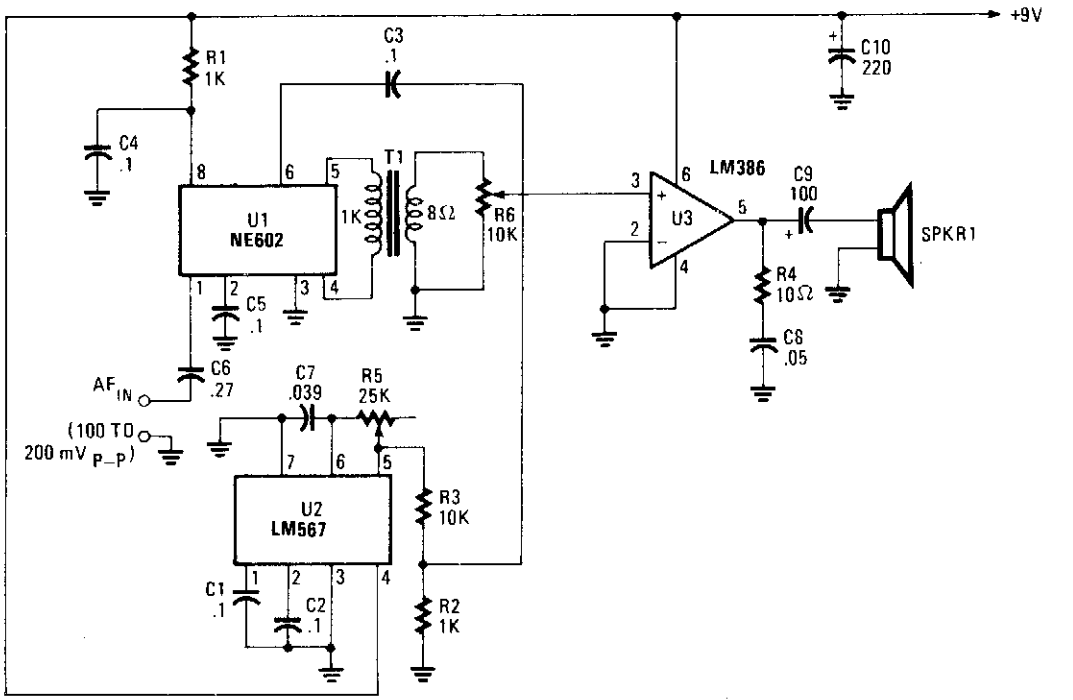
Op amp U1 operates as a unity-gain buffer to drive the input to the compressor. However, U1 is optional if a low-impedance signal source is used. The lamp's resistance will increase with temperature, which reduces the ratio of resistor R3 to the resistance of the lamp. This ratio reduces the gain of U2. The lamp's cold resistance should be greater than the input resistance of U2 (more than 50 Ω) for proper operation. The lamp's resistance will change slightly for low input levels. Therefore, the ratio of R3 to the resistance of the lamp and the gain of U2 stays high.

6

Audio Scramblers

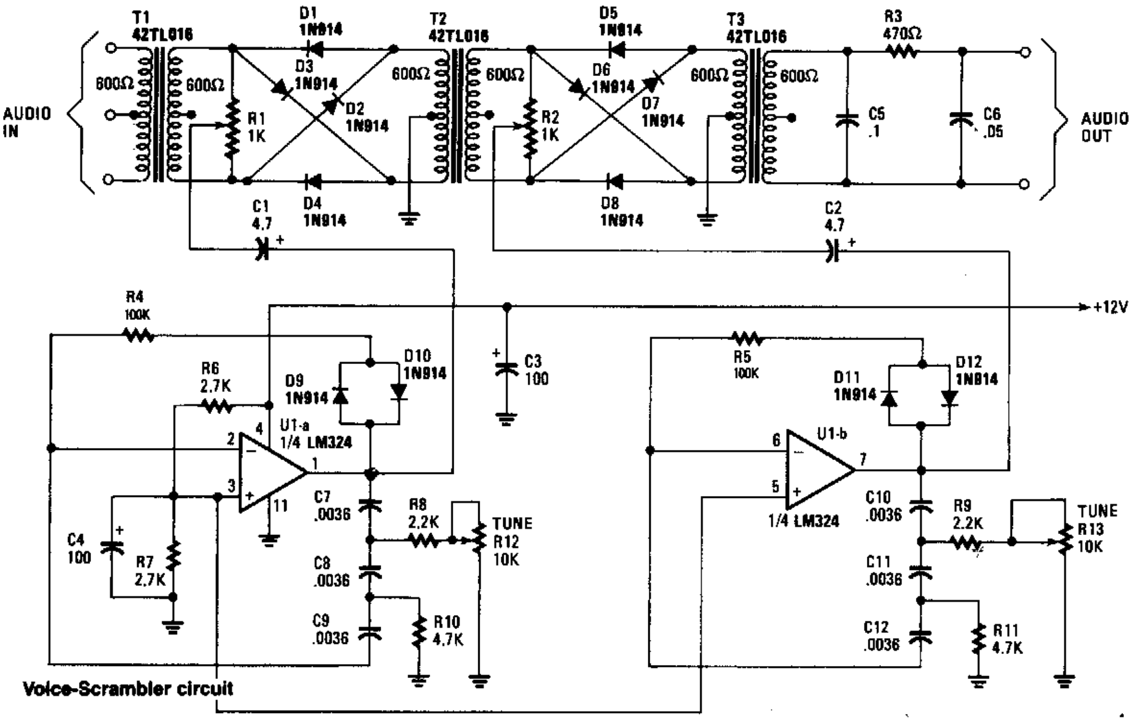
The sources of the following circuits are contained in the Sources section, which begins on page 661. The figure number in the box of each circuit correlates to the entry in the Sources section.

VOICE SCRAMBLER/DESCRAMBLER



POPULAR ELECTRONICS *Fig. 6-1*

This circuit uses an NE602 as an inversion mixer. U2 is set to run at about 2.5 to 3.5 kHz. U3 drives a loudspeaker. Because speech inversion scrambling is its own inverse, the circuit will also descramble.



Voice-Scrambler circuit

HANDS-ON ELECTRONICS *Fig. 6-2*

This circuit uses two balanced modulators to produce a DSB signal and then reinsert the carrier, except the carrier now has a different frequency. This causes an input signal to be distorted. A voice signal will be recognizable with this circuit, but the original speakers' voice will not be identifiable with correct adjustments.

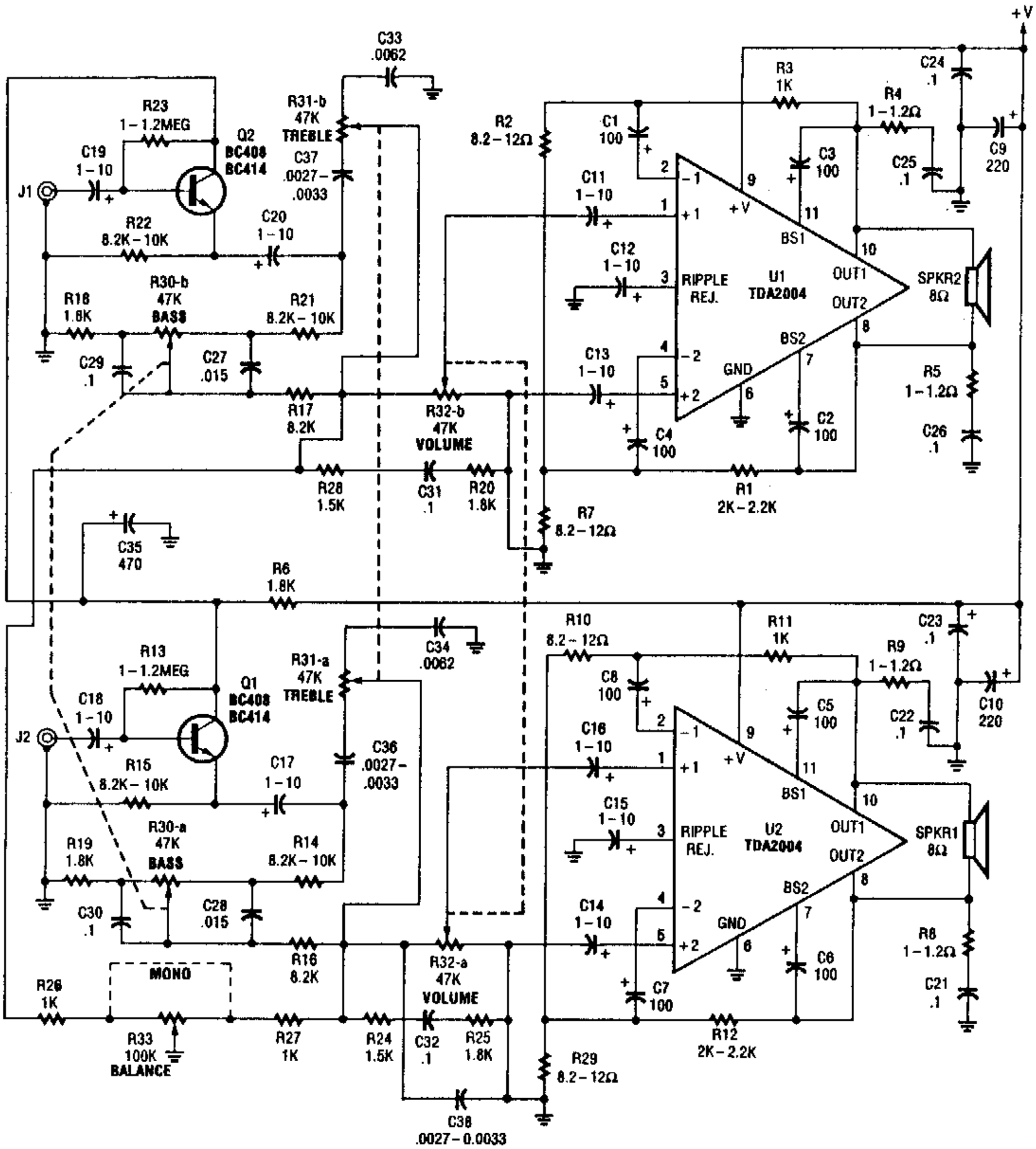
Two LM324 op amps act as oscillators that are tuneable from 2 to 3.5 kHz. The frequencies are set with R12 and R13. T1, T2, and T3 are 600 Ω CT/600 Ω audio transformers—available from Mouser Electronics, Inc.

7

Audio Power Amplifiers

The sources of the following circuits are contained in the Sources section, which begins on page 661. The figure number in the box of each circuit correlates to the entry in the Sources section.

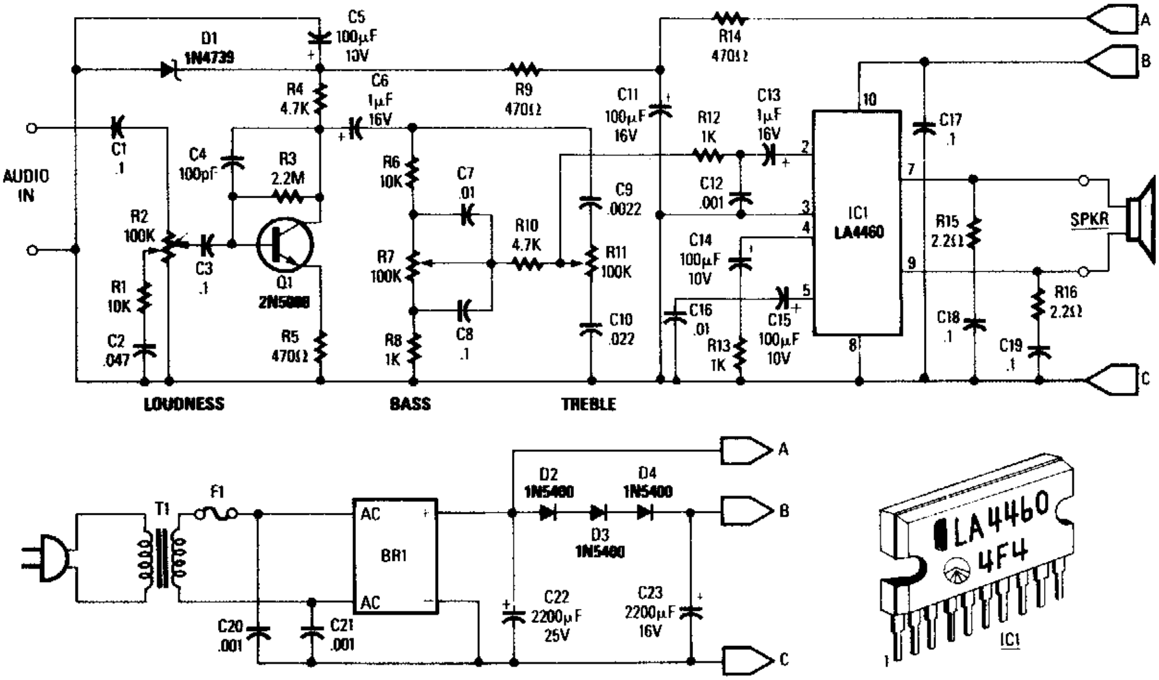
12-V/20-W STEREO AMPLIFIER



POPULAR ELECTRONICS *Fig. 7-1*

This amplifier delivers 20 W per channel. Input sensitivity is about 300 mV into 47 kΩ. Notice that a bridged output is used, so the speakers are operated with both wires above ground. A +12-V supply is used. U1 and U2 must be heatsinked.

GENERAL-PURPOSE 5-W AUDIO AMPLIFIER WITH ac POWER SUPPLY

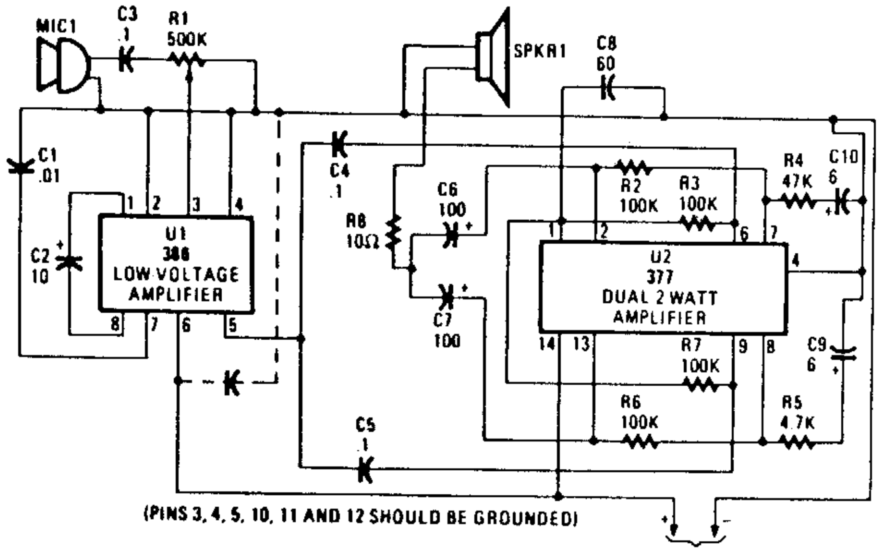


RADIO-ELECTRONICS

Fig. 7-2

This general-purpose low-power (5 W) audio amplifier is suitable for driving a speaker of approximately 8 to 12 inches. A Sanyo LA4460 IC is used as the audio output IC. The circuit consists of a loudness control, driver amplifier Q1, and bass and treble controls of about ±10 dB boost/cut. It should be useful in a wide variety of situations. Either the ac supply shown can be used, or a 12 Vdc supply can be connected to points A&B (positive) and C (negative). Two of these circuits, using ganged potentiometers at R2, R7, and R11 can be used for stereo applications. T1 is a 12-V 1-amp plug-in transformer. Notice that IC1 must be heatsinked. Power output is about 5 W. A 4″×2″×0.050″ aluminum heatsink should be adequate.

BULL HORN

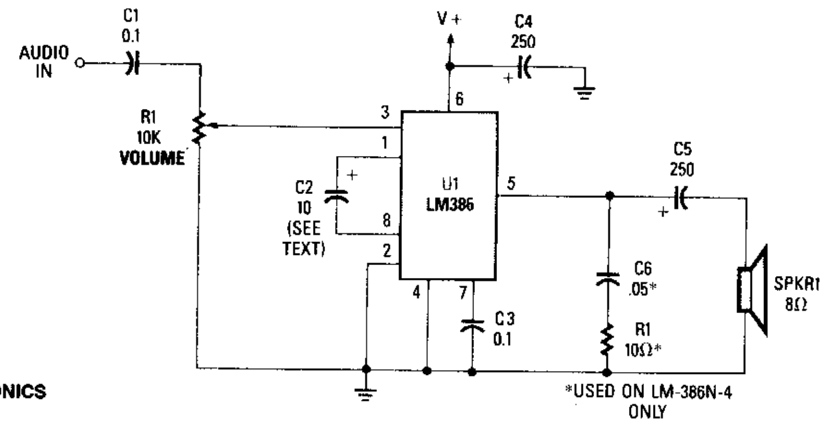


WELS' THINK TANK

Fig. 7-3

This bull horn uses U1 as a driver stage and U2 as an output driver. U1 is set up for a gain of 200. The microphone should have about 200-mVpp output. The two sections of U2 produce about 4-W of output power. Use shielded cable for all audio leads. Power is a 6-to 9-V battery or other source.

RECEIVER AUDIO CIRCUIT

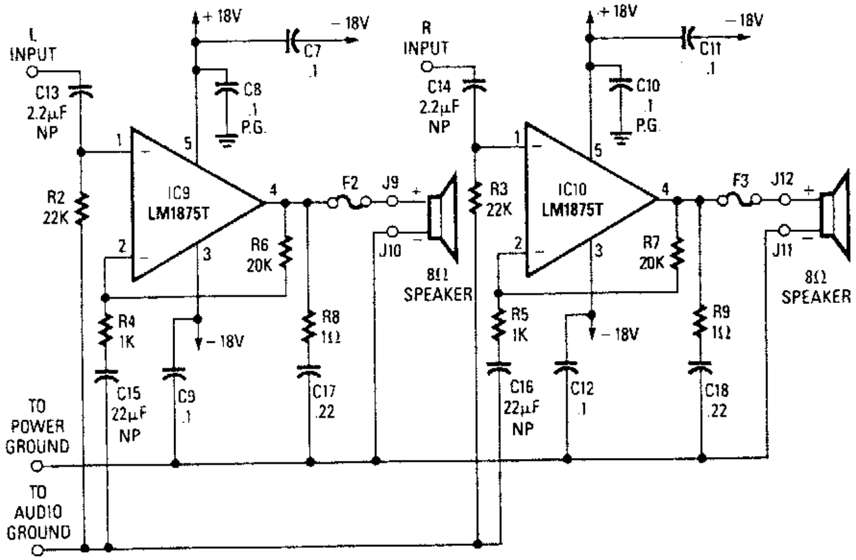


POPULAR ELECTRONICS

Fig. 7-4

This simple receiver AF amplifier can supply several hundred milliwatts to an 8-Ω speaker. The gain is about 200X. If high gain is not needed, C2 can be deleted and a gain of 20 will be obtained. R1 and C6 are musts, otherwise ultrasonic (30 to 60 kHz) oscillations might occur. C6 can be 0.1 μF on all LM386N versions for protection against these oscillations. The supply voltage is typically 6 to 12 V. No heatsink is necessary, but good grounding is a must.

AUDIO AMPLIFIER

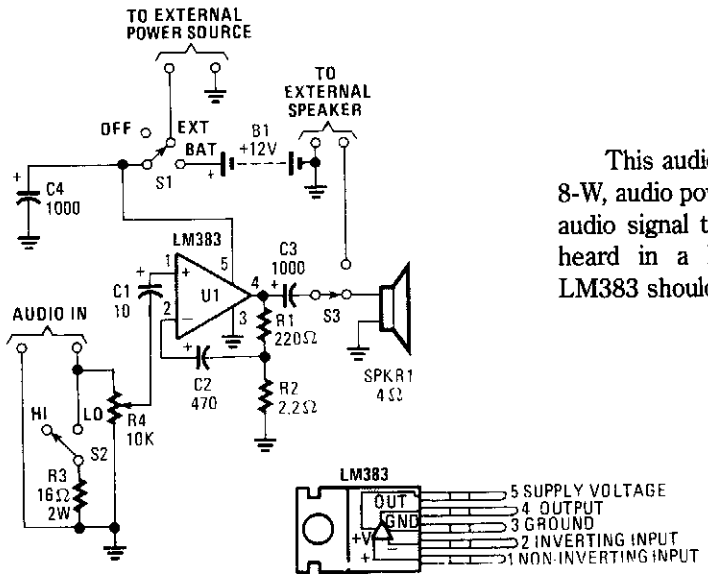


RADIO-ELECTRONICS

Fig. 7-5

This amplifier will deliver around 20 W to an 8-Ω speaker.

8-W AUDIO AMPLIFIER

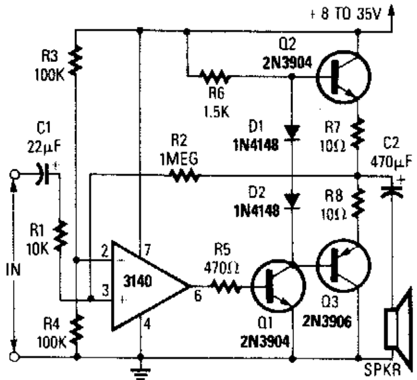


This audio power amp (built around an LM383 8-W, audio power amplifier) can be used to boost an audio signal to a sufficient level so that it can be heard in a high-noise environment. Note that LM383 should be heatsinked.

POPULAR ELECTRONICS

Fig. 7-6

SIMPLE OP AMP AUDIO AMPLIFIER



RADIO-ELECTRONICS *Fig. 7-7*

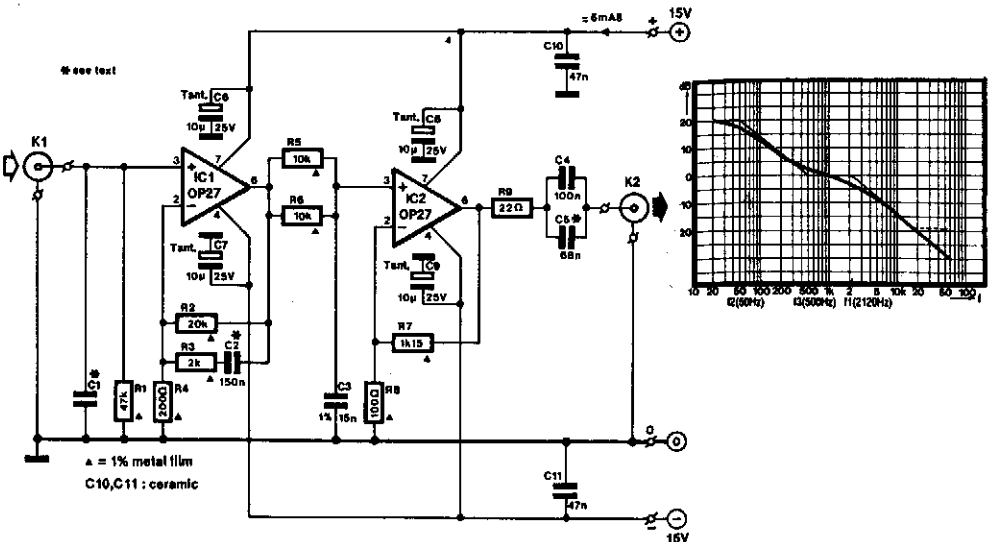
A CA3140 drives a complementary output stage Q1, Q2, and Q3. Output power depends on supply voltage and limits on dissipations of Q2 and Q3, but it can be 1 or 2 W with a higher impedance speaker and a 30-V supply.

8

Audio Signal Amplifiers

The sources of the following circuits are contained in the Sources section, which begins on page 661. The figure number in the box of each circuit correlates to the entry in the Sources section.

PREAMPLIFIER FOR MAGNETIC PHONO CARTRIDGES



ELEKTOR ELECTRONICS

Fig. 8-1

This amplifier is intended to be added to preamplifiers that have no phono input. Such a phono input is required for normal record players with a dynamic pick-up, of which millions are still around. Moreover, the amplifier does not only bring the output of the pick-up to line level, it also adds the correction to the frequency response (according to RIAA requirements).

When recording gramophone records, the frequency characteristic is lifted at the high end. This lift must be countered in the playback (pre)amplifier. The corrections to the frequency response characteristic are according to a norm set by the Record Industries Association of America (RIAA) and also by the IEC.

The corrective curve provided by the amplifier is shown in the graph (bold line). The thin line shows the ideal corrective curve. The sharp bends in this at 50 and 500 Hz are nearly obtained in the practical curve by network R3/C2; just above 2 kHz is approached in practice by filter R5/R6/C3. The arrangement of R3/C2 in the feedback loop of IC1 gives noticeably better results than the usual (passive) filter approach.

Circuit IC1 provides a dc amplification of 40 dB, which drops to about 20 dB when the frequency rises above 500 Hz. To minimize the (resistor) noise and the load of the op amp at higher frequencies, the value of R3 is a compromise. The associated polystyrene capacitor, C2, should have a tolerance of 1 to 2%.

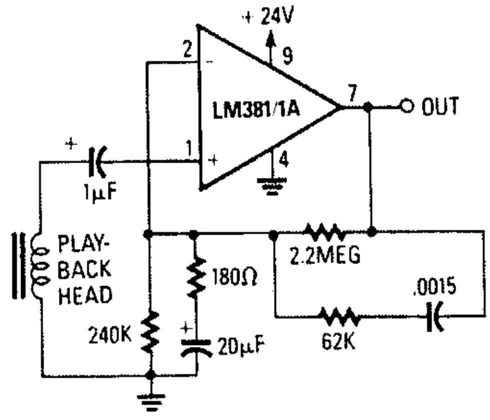
To raise the 2-mV output of the dynamic pick-up to line level at 1 kHz, linear amplifier IC2 has been added. This stage has a gain of 22 dB, so a signal of 250 mV is available at its output.

Capacitors C4/C5 at the output, in conjunction with the input impedance of the following preamplifier, form a high-pass filter with a cut-off frequency of 20 Hz; this serves to suppress any rumble or other low frequency noise. The value of C1 is normally given in the instruction booklet of the dynamic pick-up.

The power supply for the amplifier must be of good quality. Particularly, the transformer should be class A1 with a small stray magnetic field.

When the amplifier is built into the record player (best), the power supply should not be included unless it is very well screened; otherwise, hum is unavoidable.

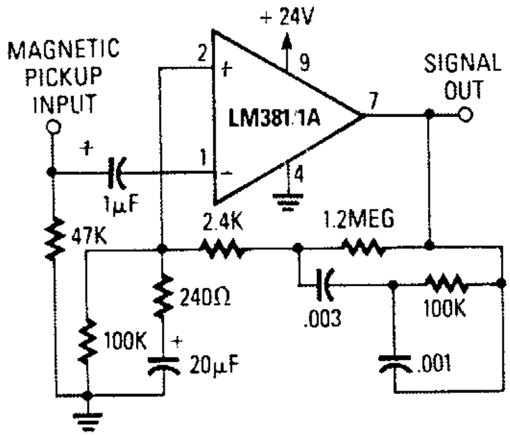
SIMPLE TAPE PLAYBACK AMPLIFIER



RADIO-ELECTRONICS *Fig. 8-2*

This circuit uses an LM381/1A as a tape preamp. The feedback network includes NAB Equalization.

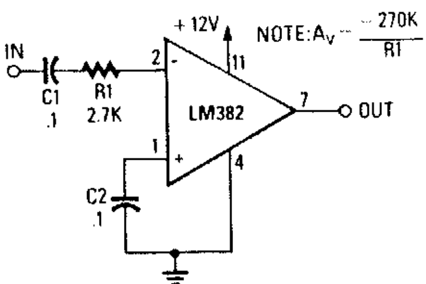
LOW-NOISE PHONO PREAMP



RADIO-ELECTRONICS *Fig. 8-3*

This circuit uses an LM381/1A as a low-noise phono preamp. The feedback network provides RIAA compensation.

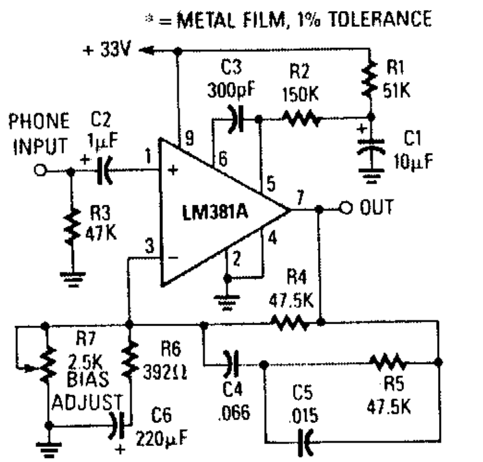
SIMPLE 40-dB GAIN AMPLIFIER



RADIO-ELECTRONICS *Fig. 8-4*

An LM382 low-noise preamp is used here to obtain a 40-dB gain amplifier, using only the IC and three peripheral components.

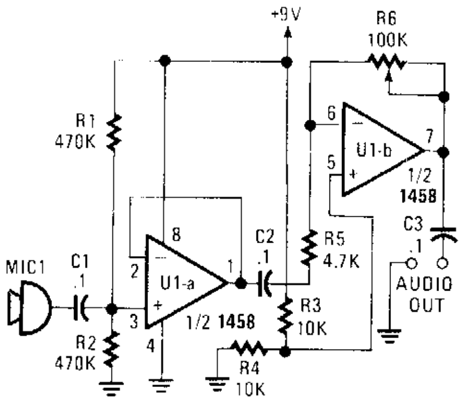
ULTRA-LOW-NOISE MAGNETIC PHONO PREAMP



RADIO-ELECTRONICS *Fig. 8-5*

This phono preamp uses an LM381/1A in a circuit that includes RIAA equalization. Adjust R7 for a voltage that is equal to half of the supply voltage (≈16.5 V).

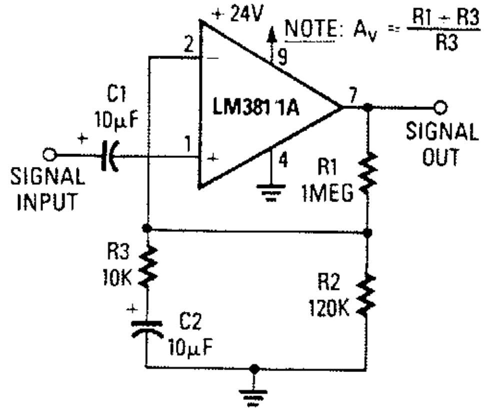
IMPEDANCE-MATCHING PREAMP



POPULAR ELECTRONICS *Fig. 8-6*

This circuit will match a crystal microphone to a device that requires a low-impedance dynamic microphone.

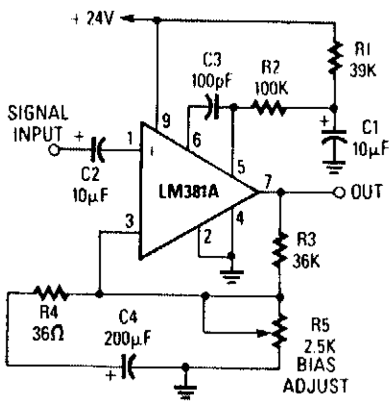
LOW-NOISE AMPLIFIER



RADIO-ELECTRONICS *Fig. 8-7*

This low-noise LM381/1A noninverting amplifier has a gain of 100.

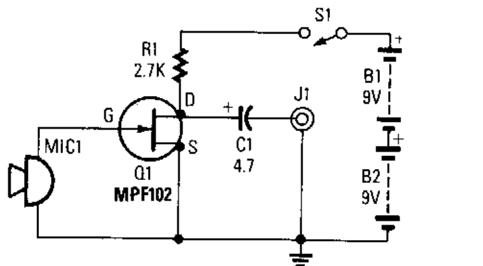
LOW-NOISE 1 000 × PREAMP



RADIO-ELECTRONICS *Fig. 8-8*

An LM381A is used here as a low-noise preamp with a gain of approximately 1 000×. Adjust R5 for 12 V at pin 7, assuming a 24-V supply.

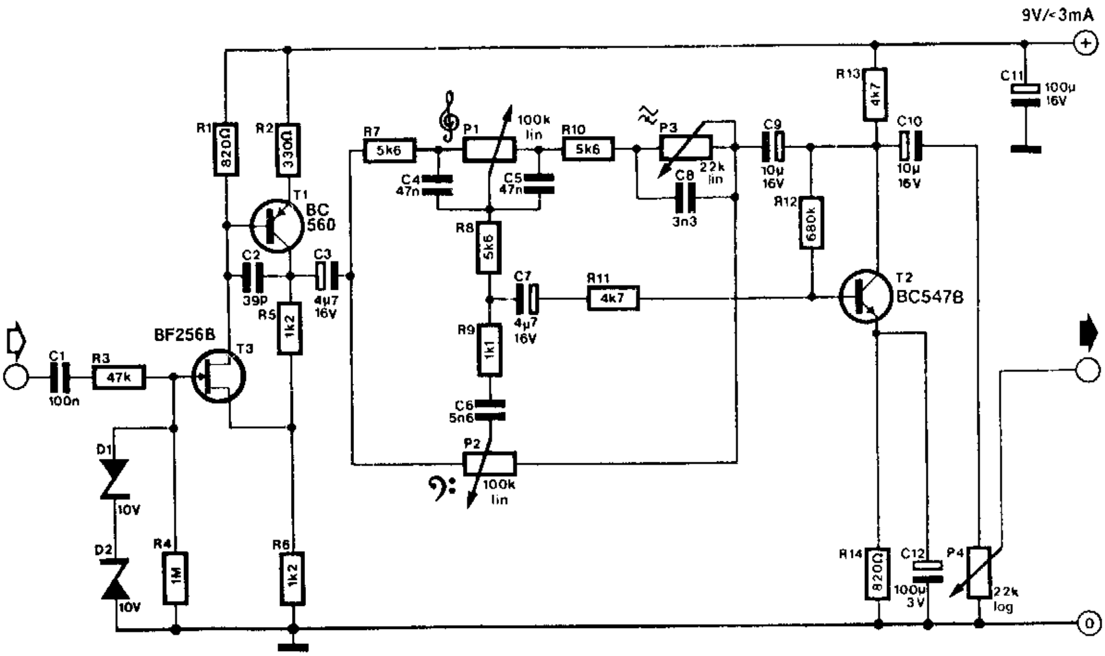
SIMPLE MICROPHONE PREAMP



POPULAR ELECTRONICS *Fig. 8-9*

This preamp uses a small dynamic microphone coupled to the gate of Q1. R1 is a load resistor. Audio is taken out between the negative side of C1 and ground. Output will be between 10 and 100 mVpp, depending on the microphone.

ELECTRIC GUITAR MATCHING AMPLIFIER



ELEKTOR ELECTRONICS

Fig. 8-10

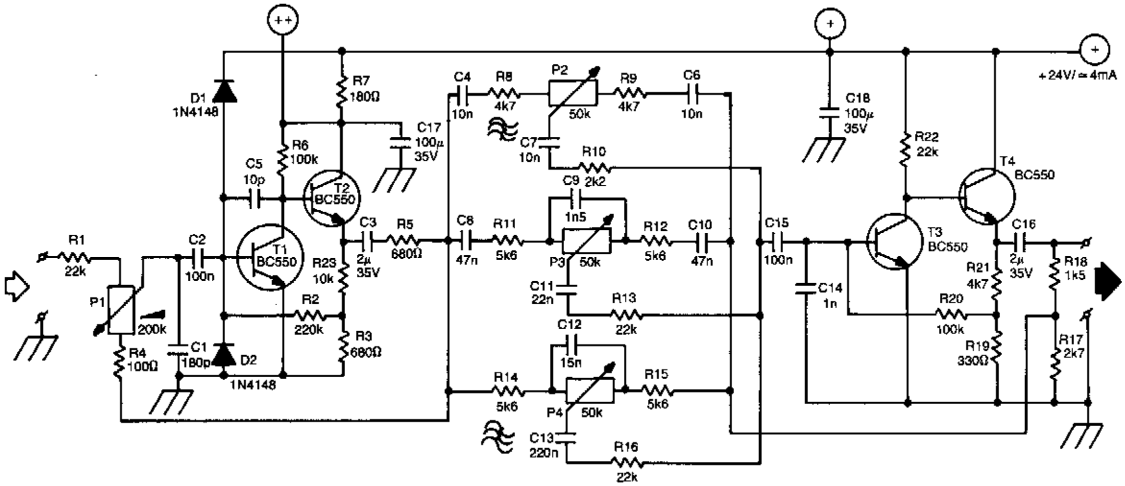
An electric guitar often has to be connected to a mixing panel, a tape deck or a portable studio. As far as cabling is concerned, that is no problem, but matching the high impedance of the guitar element to the low impedance of the line input of the mixing panel or tape deck is a problem. Even the so-called high impedance inputs of those units are not suitable for the guitar output. When the guitar is connected to such an input, hardly any signal is left for the panel or deck to process.

It would be possible to connect the guitar to the (high impedance) microphone input, but it is normally far too sensitive for that purpose; guitar clipping occurs all too readily.

The matching amplifier presented here solves those problems: it has a high-impedance (1 MΩ) input that can withstand voltages of over 200 V. The output impedance is reasonably low. Amplification is ×2 (6 dB). Dual tone control, presence control, and volume control are provided.

The circuit can handle input levels of up to 3 V. Above that level distortion increases, but that is, of course, a good thing with guitar music. Real clipping of the input signal does not occur until much higher levels than are obtainable from a guitar are applied. Power is supplied by a 9-V (PP3) battery from which the circuit draws a current that does not exceed 3 mA.

UNIVERSAL AUDIO LINE AMPLIFIER



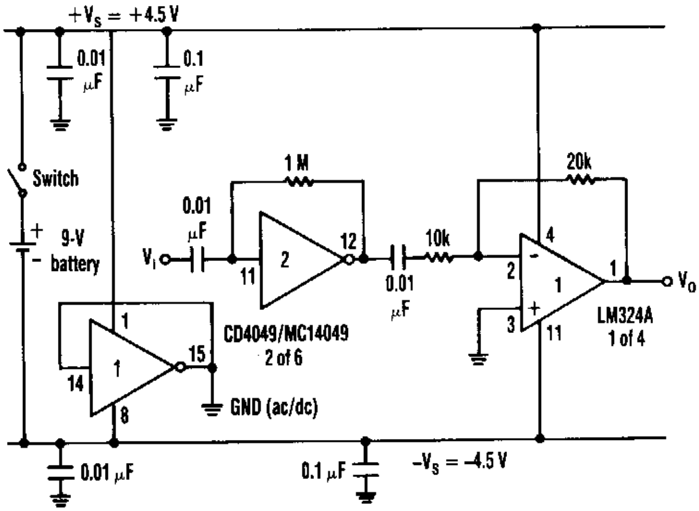
ELEKTOR ELECTRONICS *Fig. 8-11*

A line amplifier is always a useful unit to have around, whether it is for matching a line signal or raising its level somewhat. This might be needed during a recording session or with a public-address system. Furthermore, a line mixer can be constructed from a number of these amplifiers. The input of the amplifier is high-voltage proof. The output impedance is low.

The circuit is a conventional design: two dc-coupled stages of amplification are separated by a threefold Baxandall tone control system. The volume control at the input is conspicuous by having its "cold" side connected, not to ground, but to the output of the first amplifier. Because the signal there is out of phase with the input signal, the amplifier obtains negative feedback via P1. The amplification is therefore inversely proportional to the magnitude of the input signal. Thus, it is possible for the amplifier to accept a wide range of input levels. It is quite possible to input a signal taken directly from the loudspeaker terminals of a power amplifier.

The supply voltage is 24 V. At that voltage, the amplifier draws a current of about 4 mA. If several amplifiers are used in conjunction (as, for instance, in a mixer panel), the various supplies (+ and + + in the diagram) can be interlinked. Capacitors C17 and C18, and resistor R7 don't need to be duplicated in that case.

CD4049 AMPLIFIER



INVERTER CHARACTERISTICS

	V supply	
	9V	13.6V
Av	30 V/V	40 V/V
f(−3dB)	2.5 MHz	3.5 MHz nom.
Ioh	−1.25 mA	−3.0 mA min.
Iol	8.0 mA	20.0 mA min.

ELECTRONIC DESIGN

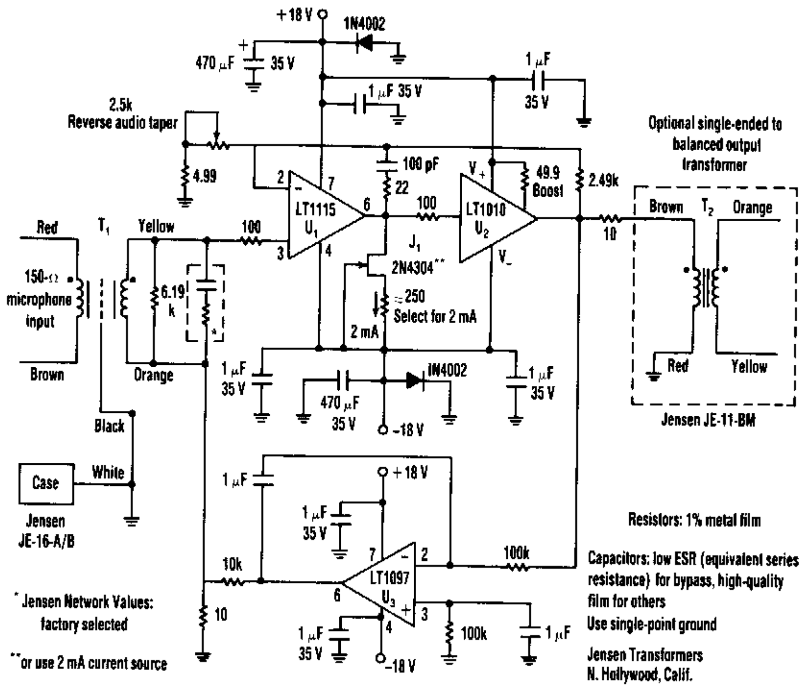
Fig. 8-12

When an inverter is biased with one resistor from its input to output in the range of 100 kΩ to 10 MΩ and is capacitor coupled, it exhibits amplifier characteristics (see the table).

Furthermore, when a split power-supply bus is needed and only one battery is used, the inverter can be configured to supply a pseudo-dc ground of relatively low impedance, coincident with the ac ground (see the figure). Depending on the magnitude of the dc ground return currents, anywhere from one inverter to several in parallel are sufficient. Also, the supply buses must be capacitor bypassed.

The configured input-to-input shorted inverter now acts as a voltage regulator that sinks and sources current. In this configuration, the inverter is forced to operate at the midpoint of its transfer characteristic. This divides the battery potential into two equal parts—as referenced to the defined dc ground by virtue of its internal gain and physical structure. Op amps such as the LM324A, can be powered from one battery while being referenced to the dc ground that is generated by the inverter. This novel technique surpasses the use of discrete resistors for battery potential dividing. It can be employed in other applications where individual component savings and improved design performance are needed.

LOW-NOISE AUDIO PREAMP

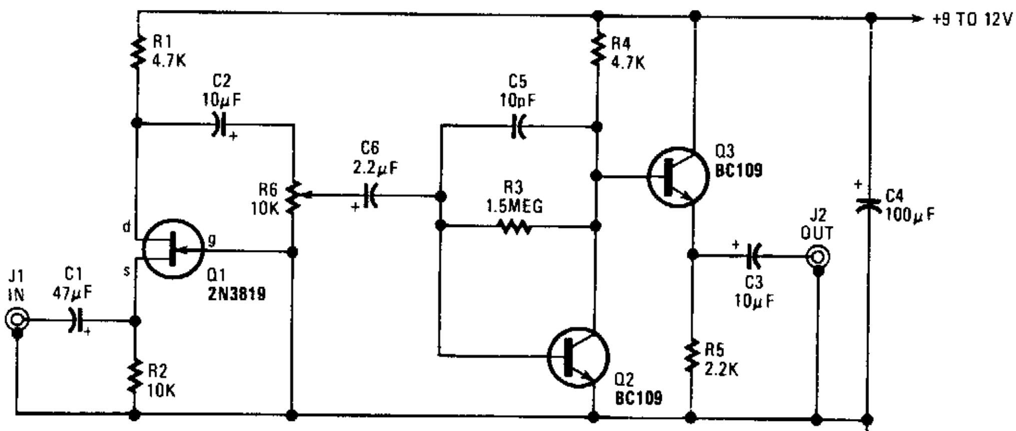


ELECTRONIC DESIGN

Fig. 8-13

A low-noise LT1115 (Linear Technology, Inc.) op amp is coupled to a class-A buffer amplifier to produce a variable gain (12-to-50 dB) microphone preamp. THD is less than 0.01% from 80 Hz to over 20 kHz. The transformers must be properly grounded and shielded.

LOW-IMPEDANCE MICROPHONE PREAMP

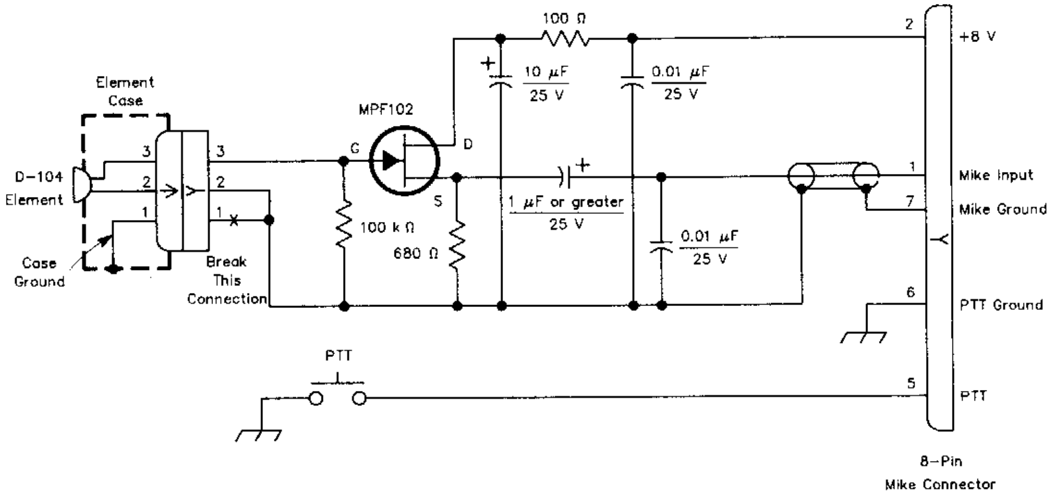


POPULAR ELECTRONICS

Fig. 8-14

This amplifier uses a common-gate FET amplifier to match a low-Z microphone.

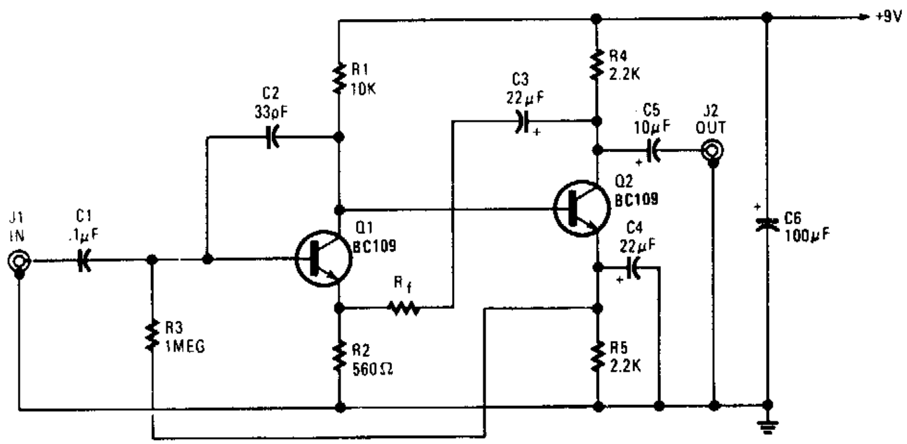
MICROPHONE PREAMP



QST *Fig. 8-15*

This circuit is used to interface a high-impedance microphone to a radio transceiver that requires a low-impedance microphone. The supply voltage can be either a battery or taken from the transceiver the circuit is used with.

GENERAL-PURPOSE PREAMP



POPULAR ELECTRONICS *Fig. 8-16*

This amplifier is useful for audio and video applications. Gain is set by R_f and the voltage gain of this amplifier is approximately $1 + R_f/560$, where R_f is in ohms. Bandwidth depends on gain selected, but typically it is several MHz. $R_f = 5.1\ k\Omega$, which produces a gain of 10× (20 dB) voltage.

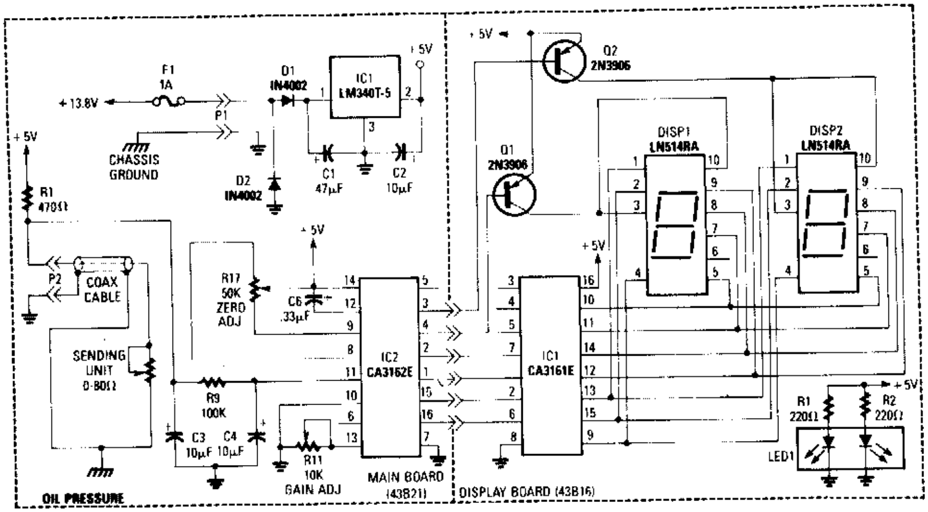
9

Automotive Instrumentation Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 661. The figure number in the box of each circuit correlates to the entry in the Sources section.

Digital Oil-Pressure Gauge
Water Temperature Gauge
Automotive Electrical Tester
Digital Vacuum Gauge
Digital Fuel Gauge
Analog Expanded-Scale Meter for Autos
Digital Pressure Gauge
Voltage Gauge
Digital Miscellaneous Temperature Gauge

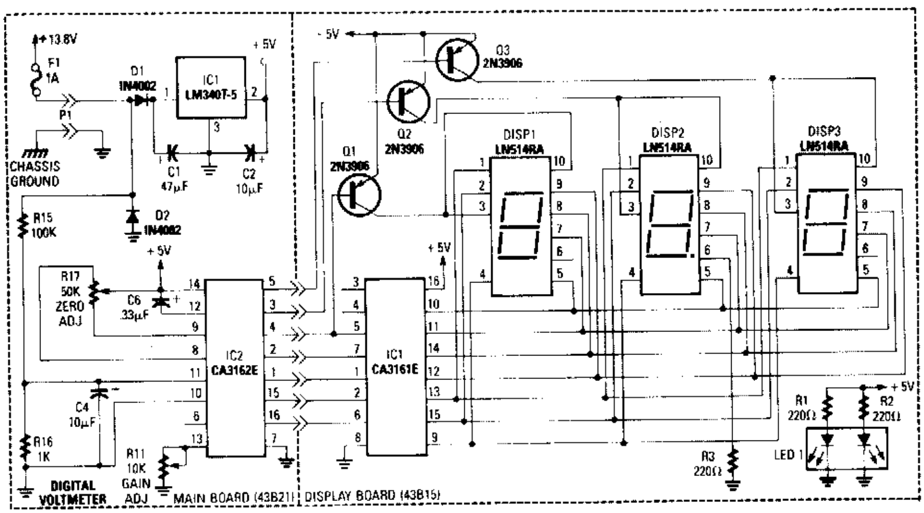
DIGITAL OIL-PRESSURE GAUGE



RADIO-ELECTRONICS *Fig. 9-1*

This gauge uses a sensor in conjunction with R1 to develop a dc voltage proportional to oil pressure. IC1 and IC2 form a two-digit DVM. Q1 and Q2 are display selectors for the multiplexed display. IC1 provides the necessary +5 V to the circuitry. Calibration is via R11 and zero adjust via R17.

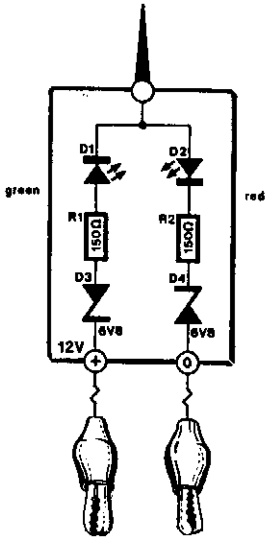
WATER TEMPERATURE GAUGE



RADIO-ELECTRONICS *Fig. 9-2*

This gauge is similar to the miscellaneous temperature gauge, except that a thermostat is used as a sensing element.

AUTOMOTIVE ELECTRICAL TESTER

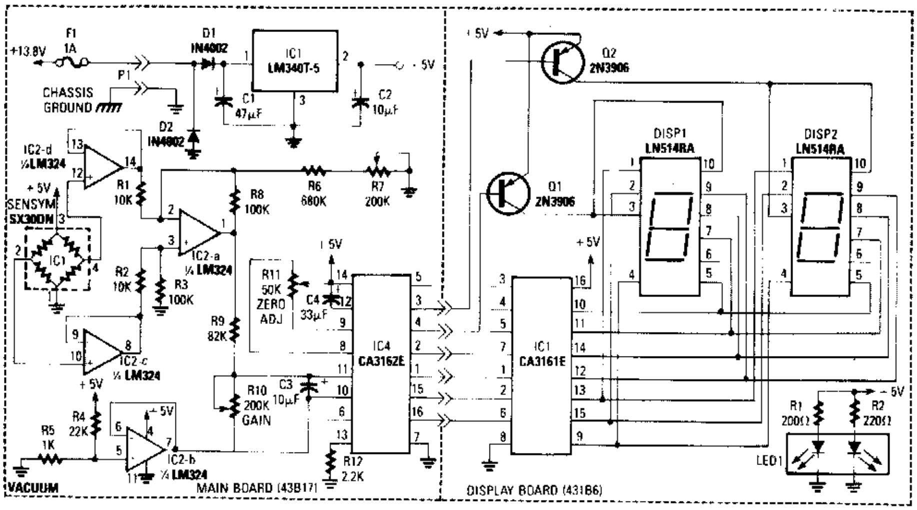


ELEKTOR ELECTRONICS *Fig. 9-3*

This little tester is useful for checking vehicle electrical circuits. Two LEDs indicate whether one of the clips is connected to the positive supply line (red) or to ground (green).

The unit is powered by the vehicle battery. It is advisable to terminate the unit into two insulated heavy-duty crocodile clips. These enable connection to be made directly to the battery or to terminals on the fuse box.It is also possible to terminate it into a suitable connector that fits into the cigarette lighter socket. If a sharp needle is soldered to one of the terminals, it is possible to check insulated wiring—but only those that carry 12 V. Although the needle pierces the insulation, it does not damage it.

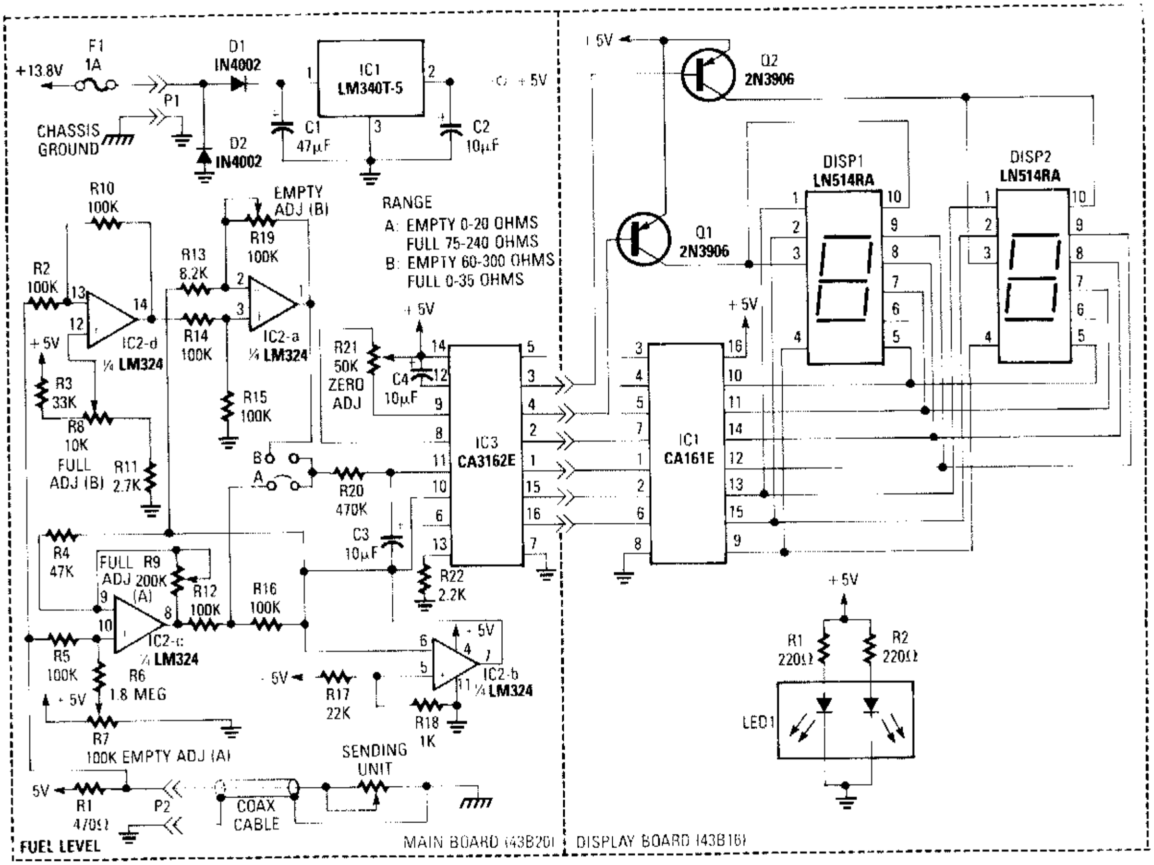
DIGITAL VACUUM GAUGE



RADIO-ELECTRONICS *Fig. 9-4*

A bridge circuit is used to produce a signal from the output of vacuum sensor IC1. IC2b provides about a 0.2 V offset for IC4, the A/D converter. IC2b and d are voltage followers that drive differential amp IC2a. The output of this circuit is used to drive IC4 and IC1, the display drivers.

DIGITAL FUEL GAUGE

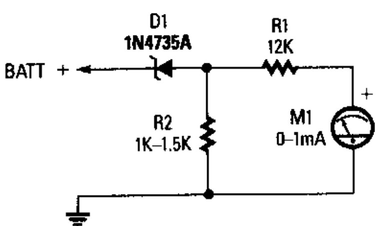


RADIO-ELECTRONICS

Fig. 9-5

This circuit uses a digital voltmeter (formed from IC1 and IC3) to display fuel quantity as a percentage of a full tank. In order to work with two kinds of fuel sensors, low resistance = full. Where higher resistance = full, IC2 forms a dc amplifier that has both inverting (path A) or noninverting (path B) outputs, and calibration adjustments for each path.

ANALOG EXPANDED-SCALE METER FOR AUTOS

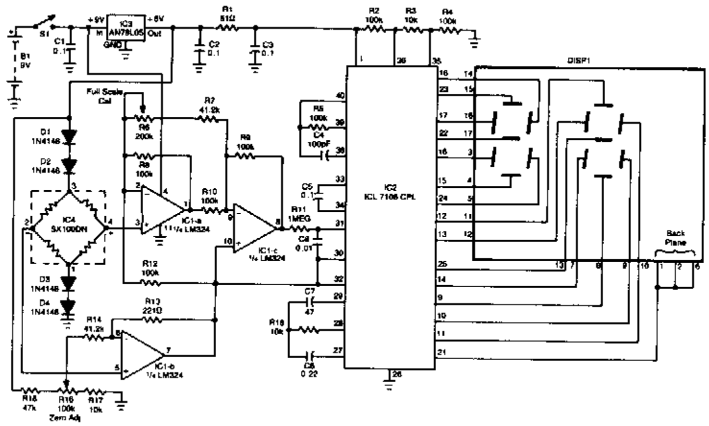


Zener diode D1 is used to suppress the first 6 V of the scale, which gives a meter reading of 6 to 8 V—useful for automotive electrical system monitoring.

POPULAR ELECTRONICS

Fig. 9-6

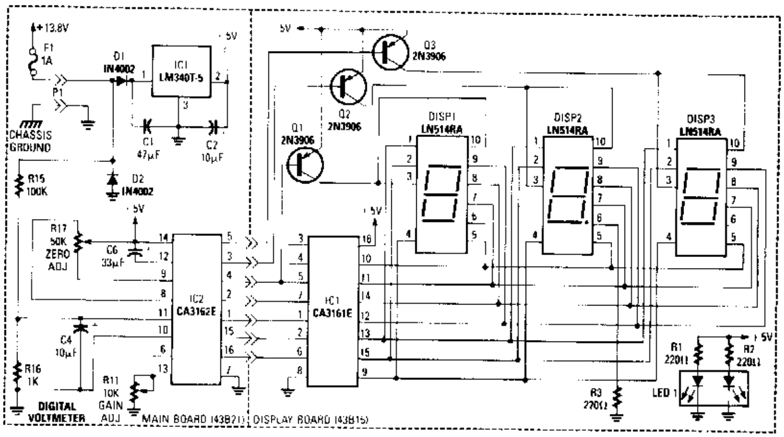
DIGITAL PRESSURE GAUGE



RADIO-ELECTRONICS *Fig. 9-7*

Using an intersil ICL 7106 A/D converter chip and an LED display module, this gauge uses a Sensym Corp. pressure transducer SX100pn (100 psi full scale) in a Wheatstone bridge configuration to drive an op amp (IC1a, b, c) translator circuit that supplies a dc voltage to IC2 that is proportional to pressure. R6 sets the gain of IC1A (full-scale sensitivity) and R16 supplies a zero adjustment. IC3 provides regulated +5 V to power the circuit.

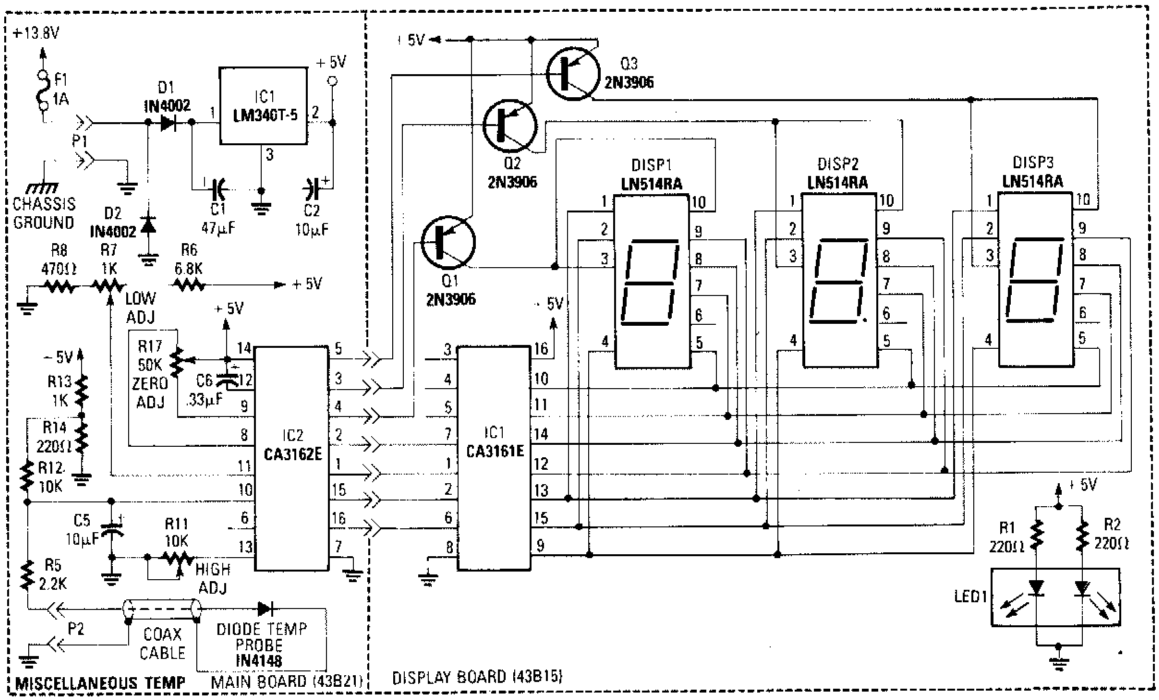
VOLTAGE GAUGE



RADIO-ELECTRONICS *Fig. 9-8*

This circuit uses an RCA CA3162E (IC2) A/D converter. This converter has 12-bit output (BCD) which is sent to display decoder driver IC1. +5 V is obtained from IC1. R17 adjusts to 0 and R11 should be set to produce correct calibration of gauge unit. Displays are common cathode types. No limiting resistors are necessary because the output drivers are constant current. R15 and R16 sample the applied voltage (usually 8 to 18 V). LED1 is used to illuminate the gauge legend (Volts, Temp, etc.).

DIGITAL MISCELLANEOUS TEMPERATURE GAUGE



RADIO-ELECTRONICS

Fig. 9-9

A diode (IN4148) is used as a temperature sensor. IC2 is an A/D converter with BCD output. A reference voltage set by R7 is applied to the positive input of IC2. As the temperature increases, the voltage across the temperature sensor resistance decreases. This increases the differential input voltage across pins 10 and 11 of IC2. R3 adjusts low calibration. R17 zeros the A/D converter and R7 adjusts high calibration.

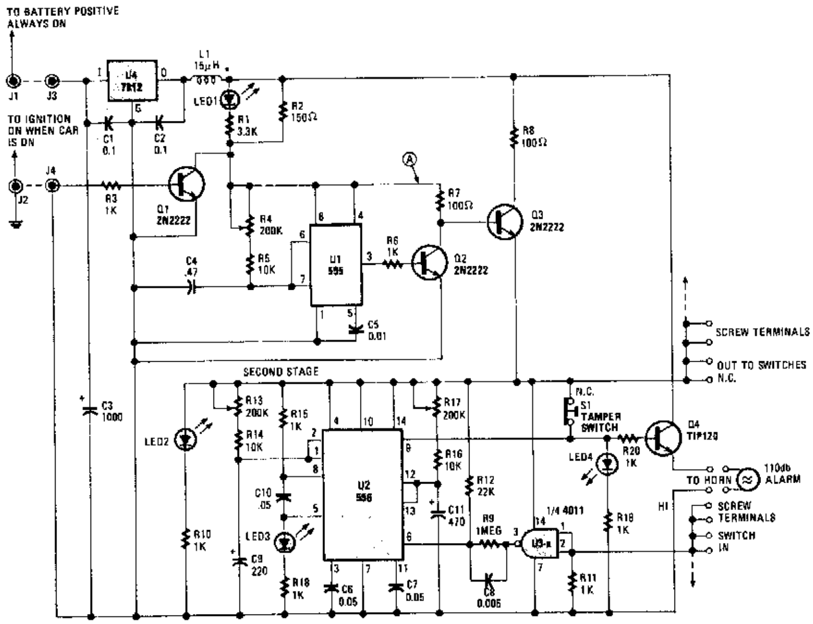
10

Automotive Security Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 662. The figure number in the box of each circuit correlates to the entry in the Sources section.

Automatic Arming Auto Alarm
Backup Beeper
Auto Turn-Off Alarm with 8-Minute Delay
Auto Alarm
Auto Ignition Cut-Off
Car Alarm with Horn as Loudspeaker
Automotic Turn-Off Alarm with Delay
Single-IC Alarm
Low-Current Simple CMOS Alarm
Back-Up Alarm

AUTOMATIC ARMING AUTO ALARM



POPULAR ELECTRONICS

Fig. 10-1

The circuit automatically turns on when the car is turned off. It gives you a variable time to get out and lock up, and also provides a variable time delay to get in and start the car.

The 555 oscillator/timers are always powered down when the car is on. That keeps the alarm from going off while you're driving. As soon as the car is turned off, Q2 switches off and shunts power to U1. When that happens, U1 immediately sends its output high, keeping Q3 on, and thereby prevents power from returning to U2.

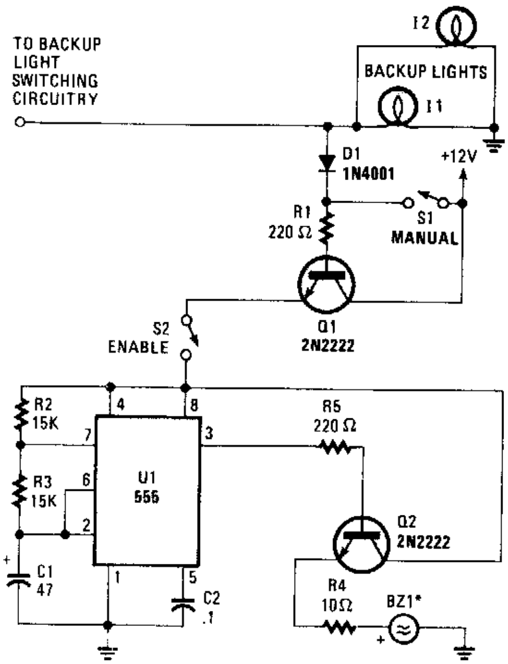
Transistor Q2 also sends power to Q3's collector to be used only when U1 has completed its timing cycle. When U1 has finished, it turns Q3 off, which in turn activates Q4, and sends power to the balance of the circuit. That timing period was the time needed to get out of the car. LED1 indicates that the system is disarmed and LED2 indicates that the system is armed.

At this point, U2 waits for a trigger pulse from the car's door switches or dome light. A positive impulse at the 4011's input sends a negative trigger pulse to the first stages of U2, which is connected as a cascading timer. The first stage's output becomes high for a time to allow the car to be turned on.

If that does not happen, the first stage's output lowers, which sends a low trigger pulse to the second stage. The second stage then sends its output high, turning on Q5, which sounds the alarm for a given time. Once that time has elapsed, the alarm is shut off by a low output to Q5 and the system is reset. If the car door is closed or a second door opened while the alarm is sounding, the first stage retriggers and prepares to extend the ON-time of the alarm.

The cascading or counting action continues until the car is left alone. You can add a switch on the positive supply rail at J3 to override and silence the alarm, if (for example) you plan to work on the car. Switch S1 is a normally closed type that is built into the case of the alarm; S1 is pushed to the open position when the case is mounted flush with a surface. Any attempt to remove the alarm will sound the alarm.

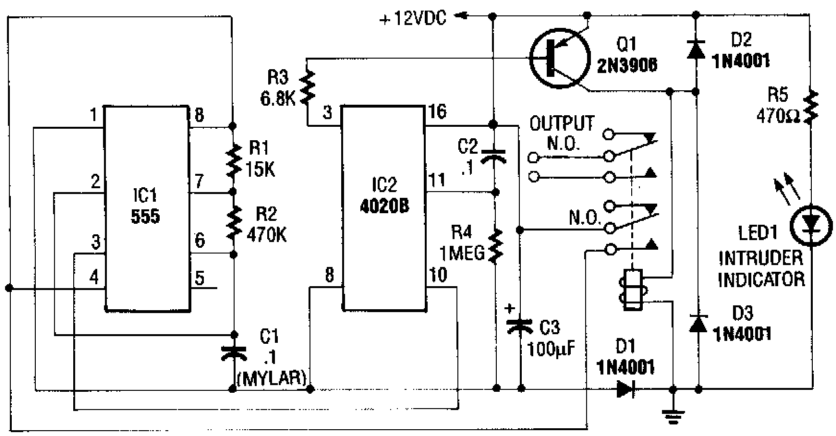
BACKUP BEEPER



POPULAR ELECTRONICS *Fig. 10-2*

When the vehicle's backup lights kick on, or when the manual switch (S1) is closed, a small current is fed to the base of Q1. Transistor Q1 allows current to flow through it and, if the enable switch (S2) is closed, it sends 12 V to U1, a 555 timer. Timer U1 sends high pulses that last 0.977 13 s and low signals that last 0.488 565 s to the base of Q2. When U1 switches Q2 on, it sends 12 V to BZ1, a piezoelectric buzzer. For best results, the buzzer should be mounted under the vehicle—somewhere where people around the car can hear the warning beeps.

AUTO TURN-OFF ALARM WITH 8-MINUTE DELAY

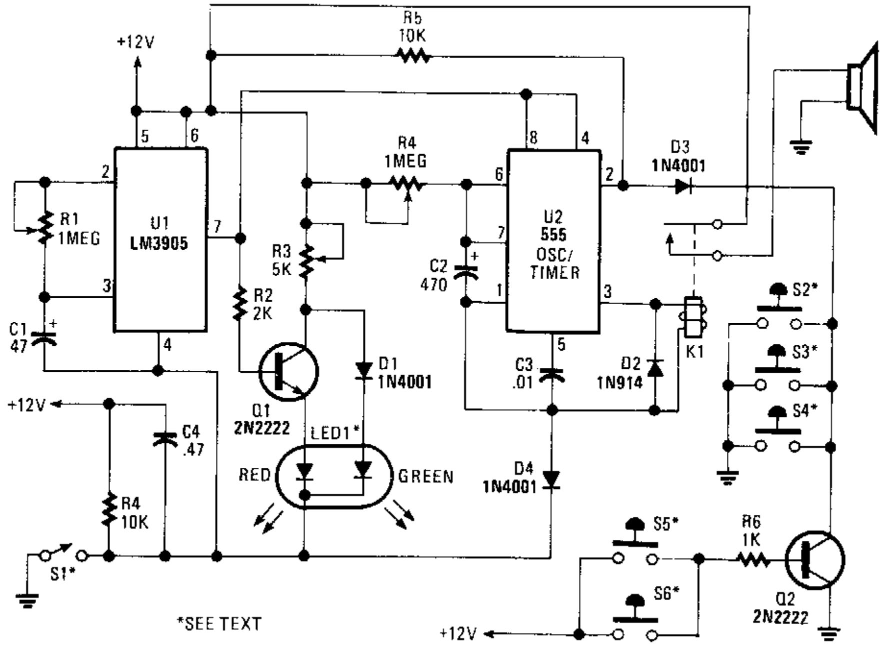


RADIO-ELECTRONICS

Fig. 10-3

This circuit uses a NE555 timer and CD4020B. When +12 Vdc is applied to the circuit, the output of IC2 is set low via C2, which turns on the relay, and IC1, a pulse generator. IC1 pulses counter IC2. After 8192 clocks, IC2 output (pin 3) goes high, cuts off Q2, and completes the cycle.

AUTO ALARM

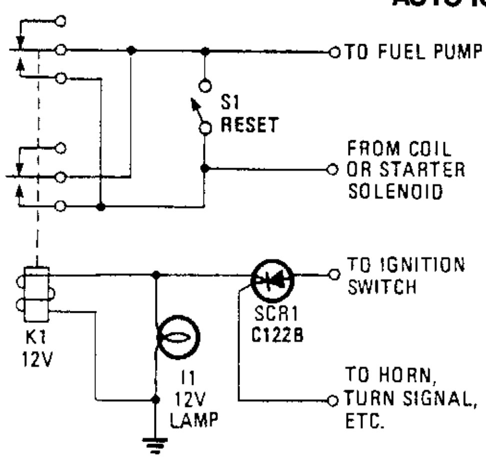


POPULAR ELECTRONICS

Fig. 10-4

S1 is external key switch. The alarm allows a 0- to 45-s delay after S1 is operated before the circuit is armed. During this period, LED1 lights green. After this delay, LED1 lights red, which indicates that the circuit is armed. Then, sensors S2 through S4 -(NO) or S5 through S6 (NO) pull pin 2 of U2 low, which activates K1 and sounds the alarm. The alarm sounds for a duration determined by R4 and C2. After this time, K1 releases and the circuit is again ready. Manual reset is via the key switch, S1.

AUTO IGNITION CUT-OFF

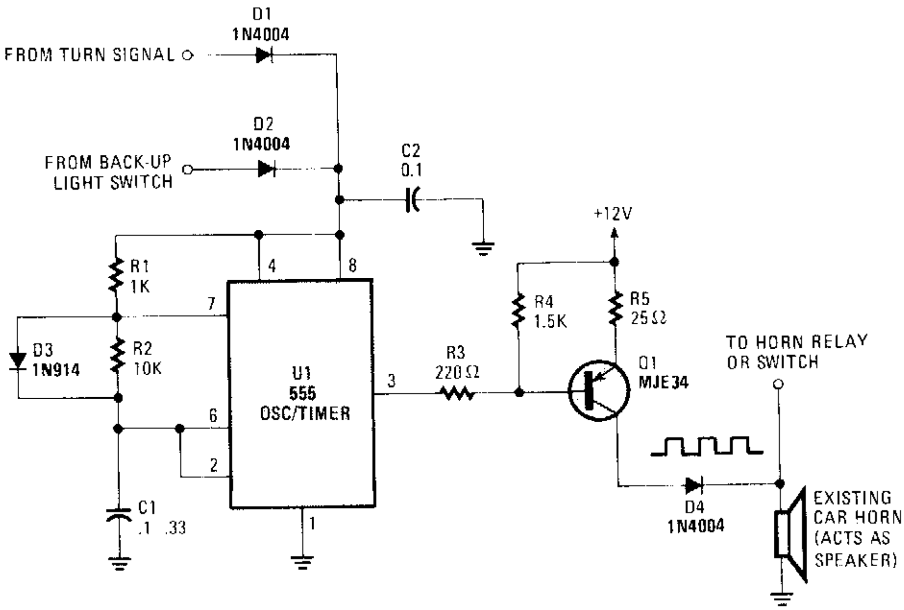


POPULAR ELECTRONICS

Fig. 10-5

Using an SCR/relay combination, this circuit can be made to cut off ignition, unless a positive voltage is applied to the gate of the SCR. This is useful as an anti-theft device, because depending on hook-up, the car will not start unless a certain accessory or a hidden switch is closed.

CAR ALARM WITH HORN AS LOUDSPEAKER

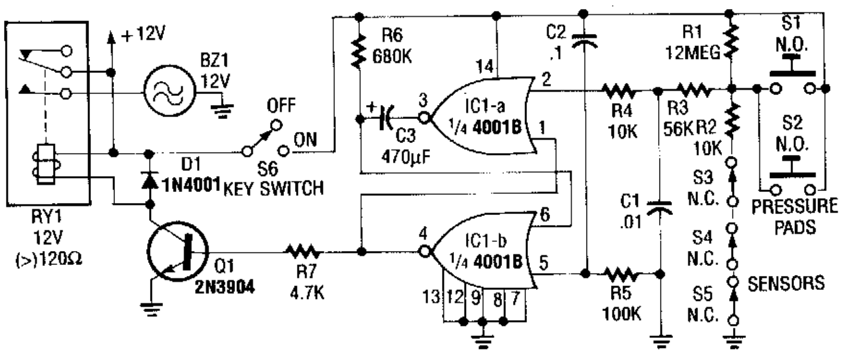


POPULAR ELECTRONICS

Fig. 10-6

An auto horn will work as a speaker of limited audio-frequency range. This circuit uses a 555 timer as an oscillator to drive an MJE34 transistor, which in turn drives the horn. Normal horn operation is ensured by blocking diode D4.

AUTOMATIC TURN-OFF ALARM WITH DELAY

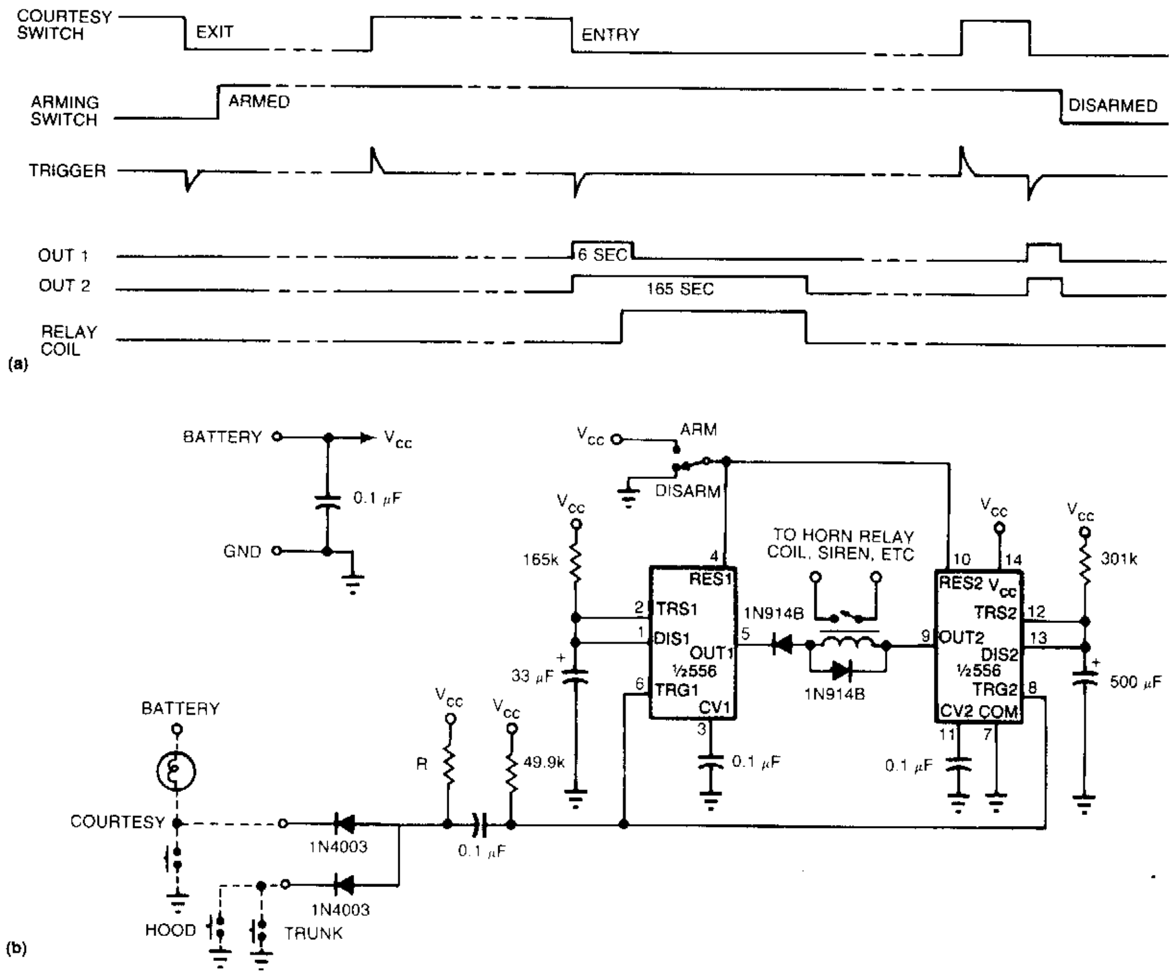


RADIO-ELECTRONICS

Fig. 10-7

In this circuit, IC1A and IC1B act as a monostable multivibrator. Any input from the sensors S1 through S5 forces IC1A to produce logic low, which causes IC1B to turn on Q1 until C3 changes through R6. This action resets the latch formed by IC1A and IC1B.

SINGLE-IC ALARM



EDN *Fig. 10-8*

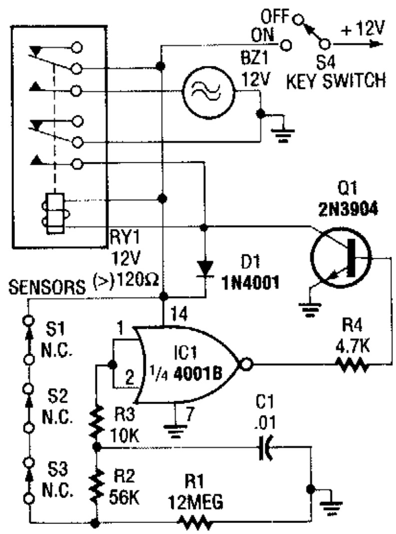
With a single IC, you can build a simple, reliable auto burglar alarm or a similar alarm. See (a) for the timing information for the alarm circuit in (b).

When you leave your vehicle, flip the arming switch and close the door behind you to arm the device. Subsequent opening of an entrance triggers both timers. After the expiration of the entry delay timer, the alarm sounds for a time that is determined by the second timer.

The value of R should be less than 1 kΩ. If you use an incandescent lamp instead of a resistor, you get an extra function—an open-entrance indicator. By keeping the resistance low, you avoid false tripping should water collect under the hood.

If your door switch connects the courtesy light to 12 V rather than to ground, use a single transistor as an inverter at the input. Although this circuit's simplicity has its drawbacks, you can add more features, such as no-entry delays for the hood and trunk, and retripping when doors remain open.

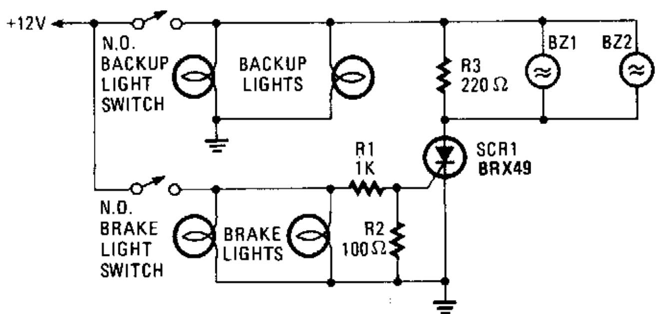
LOW-CURRENT SIMPLE CMOS ALARM



RADIO-ELECTRONICS *Fig. 10-9*

This CMOS-aided alarm draws only 1 μA standby current. An open sensor allows IC1 to bias Q1 on, activating RY1.

BACK-UP ALARM



POPULAR ELECTRONICS *Fig. 10-10*

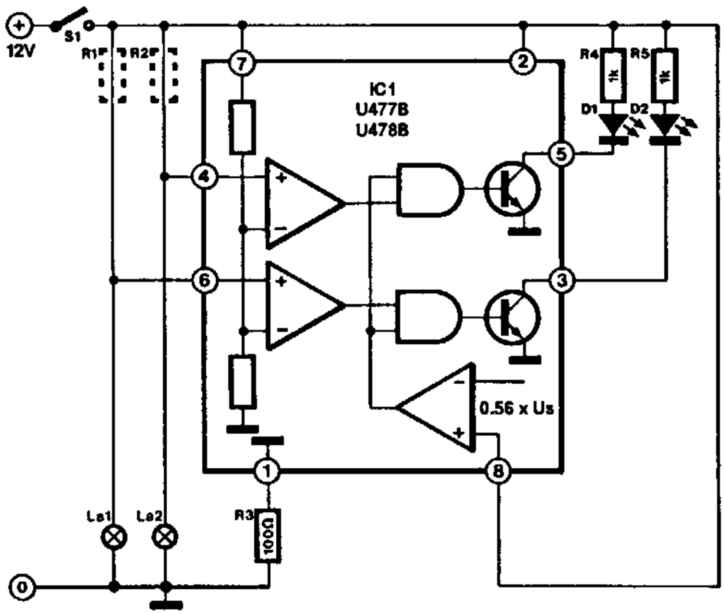
The brake lights of the automobile trigger this circuit on and off. This saves the annoyance of the alarm when it is not needed.

11

Automotive Light Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 662. The figure number in the box of each circuit correlates to the entry in the Sources section.

CAR LIGHTS MONITOR



ELEKTOR ELECTRONICS *Fig. 11-1*

This circuit is for the purpose of monitoring automotive lighting. Two special ICs are available from Telefunken that are designed to measure the current through a light bulb. In practice, detecting whether a current flows through a bulb or not is a most suitable way of determining whether the bulb still works.

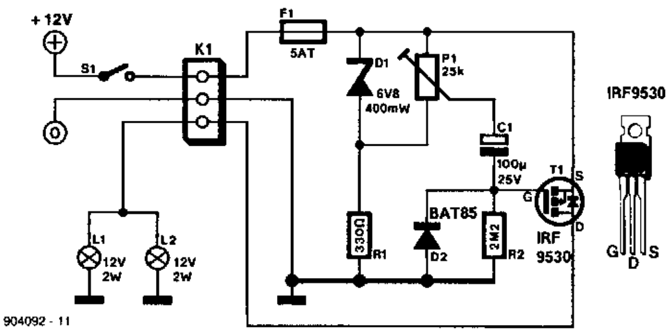
If a small resistance is connected in series with a bulb, a small voltage drop will develop across it when the bulb lights (R1 and R2 in the diagram). Each IC can cope with only two bulbs, so three or four ICs are needed per car. The junction of the bulb and resistor is connected to one of the inputs (pin 4 or pin 6) of the IC. The potential across the resistor is compared in the IC with an internal reference voltage. Depending on which of the two ICs is used, the voltage drop must be about 16 mV (U477B) or 100 mV (U478B). This voltage drop is so small that it will not affect the brightness of the relevant bulb.

The value of the series resistor is determined quite easily. If, for instance, it is in series with the brake light (normally 21 W), the current through the bulb, assuming that the vehicle has a 12-V battery, is $21 \div 12 = 1.75$ A. The resistance must then be of $16 \div 1.75 = 9$ mΩ (U477B) or $100 \div 1.75 = 57$ mΩ (U478B).

These resistors can be made from a length of resistance wire (available from most electrical retailers). Failing that, standard circuit wire of 0.7 mm diameter can be used. This has a specific resistance of about 100 mΩ per meter. However, in most cars, the existing wiring will have sufficient resistance to serve as series resistor.

LEDs can be connected to the outputs of the IC (pins 3 and 5). These will only light if the relevant car light fails to work properly.

REAR FOG LIGHT CONTROLLER WITH DELAY



ELEKTOR ELECTRONICS

Fig. 11-2

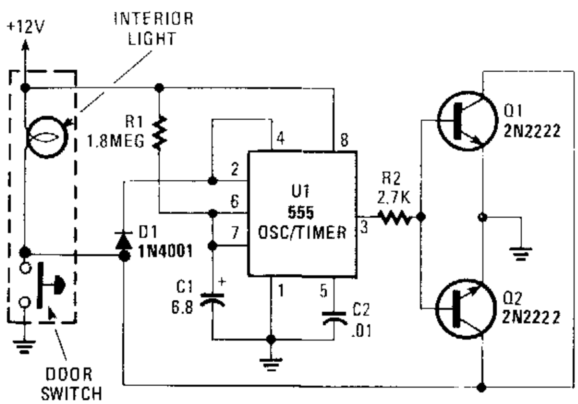
We assume that most of our readers are thoughtful drivers who do not switch on their rear fog lights when closely followed by other traffic. Following drivers (for an instant) will think that you are braking (although they have seen no reason for your doing so) and thus slam on their brakes as well. This could cause a very dangerous situation. Avoid a potentially dangerous action and install the rear fog light delay circuit, presented here.

Switch S1 is the on/off control for rear fog lights L1 and L2. As soon as this switch is closed, the gate-source voltage (Vgs) of MOSFET T1 will become more and more negative. Thus, the IC will conduct harder and harder. This in turn causes the brightness of the lights to gradually become brighter. Maximum brightness is reached after a delay of about 20 seconds, which is determined by time constant R2/C1.

The gate of T1 can be given a bias by preset P1. This provides compensation for the initial period after the lights are switched on and the lamps do not light, because they need hundreds of milliamperes before they can do so. With P1 set correctly, the lamps will light, albeit weakly, and immediately the control switch is closed. The gate potential is then equal to the voltage at the wiper of P1 (remember that C1 is then still discharged).

Although the dissipation of T1 is a maximum during the transitional period (between switch on and the lamps lighting brightly), the heatsink required is calculated on the basis of the dissipation when the lamps light brightly.

INTERIOR CONVENIENCE LIGHT

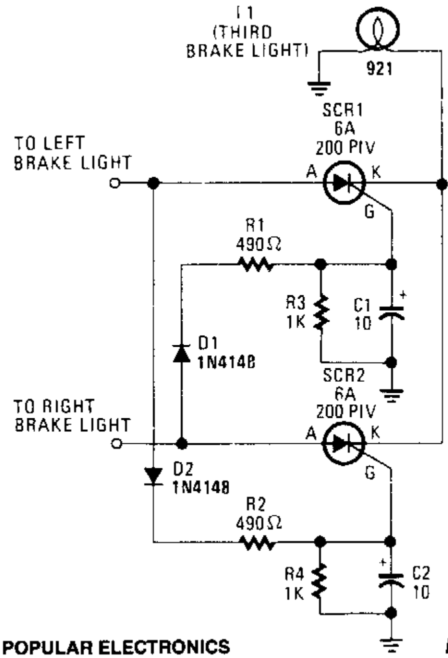


POPULAR ELECTRONICS

Fig. 11-3

When the door is closed, this circuit keeps the dome light on for a period determined by R1/C1. The time is approximately $(1.1)R_1 \times C_1$.

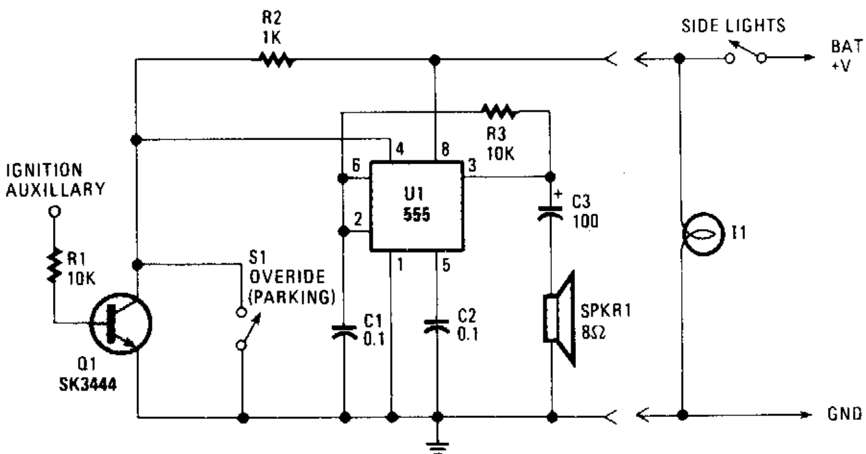
THIRD BRAKE LIGHT



POPULAR ELECTRONICS *Fig. 11-4*

The circuit is designed to light (via SCR1 and SCR2) only when both the left and right brake lights are activated. The circuit operates on 12-V negative-ground systems. When the brake pedal is depressed, 12 V is applied to the left and right brake lights. The gates of the two SCRs are triggered and current flows through the SCRs, which turns on the third brake light.

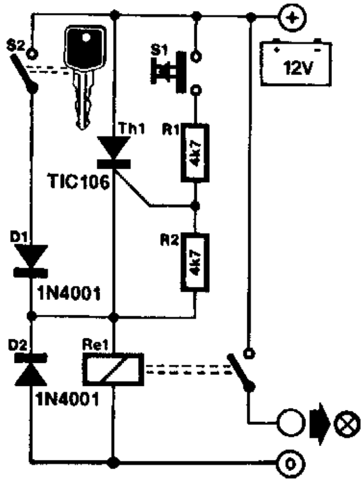
LIGHTS-ON WARNING



HANDS-ON ELECTRONICS/POPULAR ELECTRONICS *Fig. 11-5*

Because power for the circuit is obtained from the car's side lights, the circuit can't oscillate unless the lights are on. The reset pin on the 555 connects to transistor Q1. The base of Q1 is connected through R1 to the ignition auxiliary terminal on the car's fuse box. When the ignition is turned on, power is supplied to the base of Q1, which turns it on. With Q1 turned on, pin 4 of U1 is tied low, which disables the oscillator and inhibits the alarm. If the ignition is turned off while the lights are on, power is applied to the 555 and Q1 is turned off, and the alarm starts. Switch S1 is an optional override.

CAR HEADLIGHT CONTROL

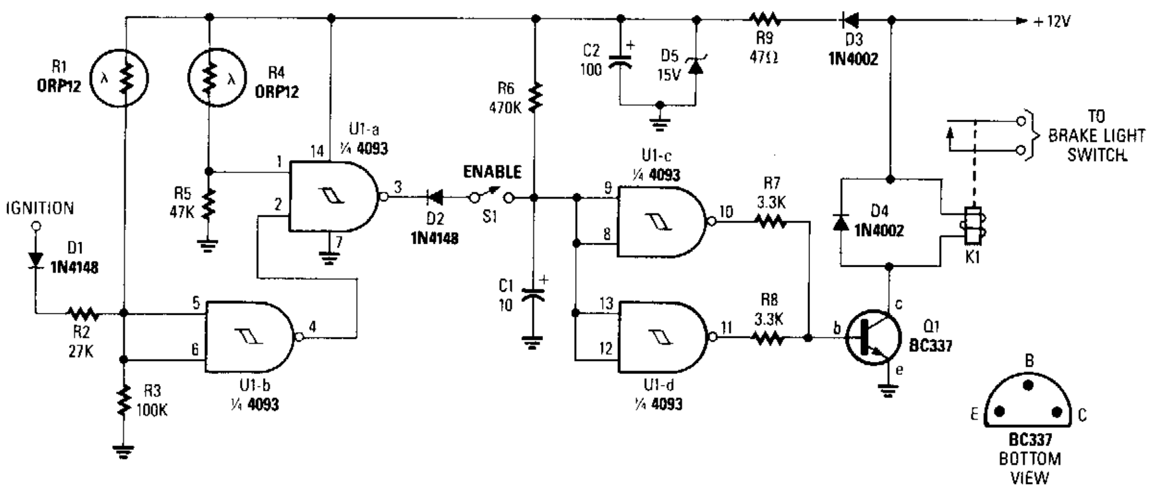


ELEKTOR ELECTRONICS *Fig. 11-6*

It is annoying to realize that you have left your car headlights on only to find that the battery is flat. One possible way to prevent this is with the present control.

The circuit does not provide a warning, but an action: when you switch off the ignition, relay Re1 is de-energized and the headlights are switched off—unless you deliberately decide otherwise. That decision is made possible by switch S1, which, when operated, triggers silicon-controlled rectifier Th1 so that Re1 is energized. Notice that this is possible only when the ignition switch, S2, is off. Otherwise, the voltage across Th1 is so low, owing to shunt diode D1, that it cannot be triggered. However, the headlights should not normally be switched on when the ignition is off; in most cases S1 will be used only rarely and the switch can then be omitted altogether. The relay should be a standard 12-V car type with contacts that can switch up to 25 A.

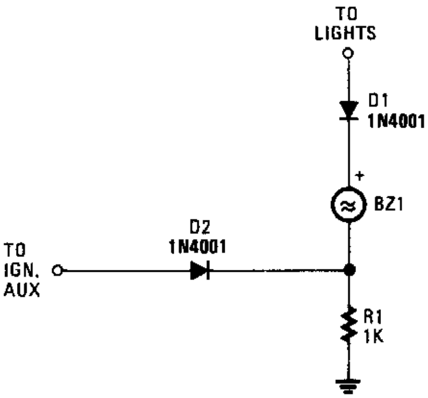
NIGHT SAFETY LIGHT FOR AUTOS



POPULAR ELECTRONICS *Fig. 11-7*

This circuit turns on the brake lights of a parked car when the headlights of an oncoming car are detected, warning the driver of the oncoming car about the parked vehicle. LDR4 is the sensor. LDR1 disables the circuit by causing U1 gate input to be pulled high during daylight hours, causing pin 2 of U1a to become low, disabling it and the circuit.

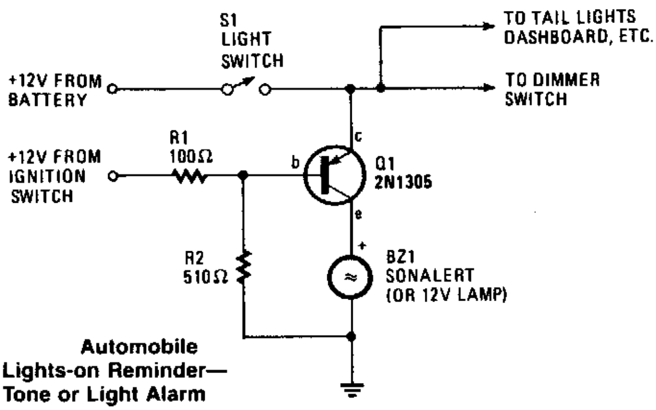
LIGHTS-ON REMINDER FOR AUTOS



POPULAR ELECTRONICS *Fig. 11-8*

When ignition is off, BZ1 will sound if the headlights are on. With the ignition on, BZ1 receives no voltage.

AUTOMOBILE LIGHTS-ON REMINDER



Automobile Lights-on Reminder— Tone or Light Alarm

HANDS-ON ELECTRONICS *Fig. 11-9*

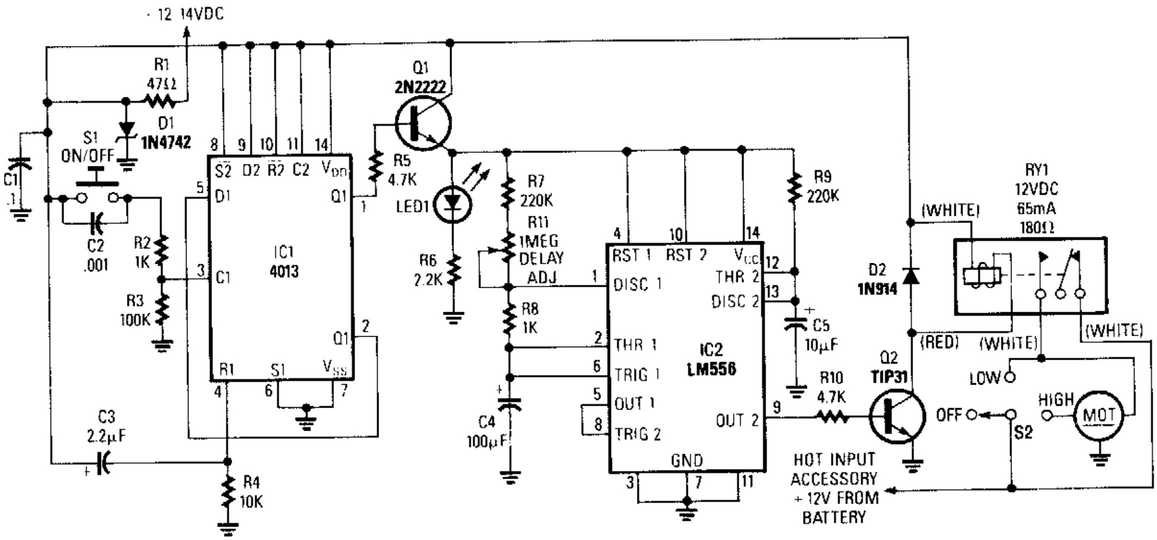
The circuit can be used to give a visible or an audible warning that the headlights are on. It uses a 2N1305 transistor as a switch to turn on a Sonalert tone generator or a small 12-V lamp. Operating current for the transistor is supplied from the wire that feeds the headlights. When the ignition is on, the transistor is biased off and the alarm is not activated. Turning off the ignition while the lights are on sets off the alarm.

12

Automotive Electronic Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 662. The figure number in the box of each circuit correlates to the entry in the Sources section.

SOLID-STATE WINDSHIELD-WIPER DELAY



RADIO-ELECTRONICS

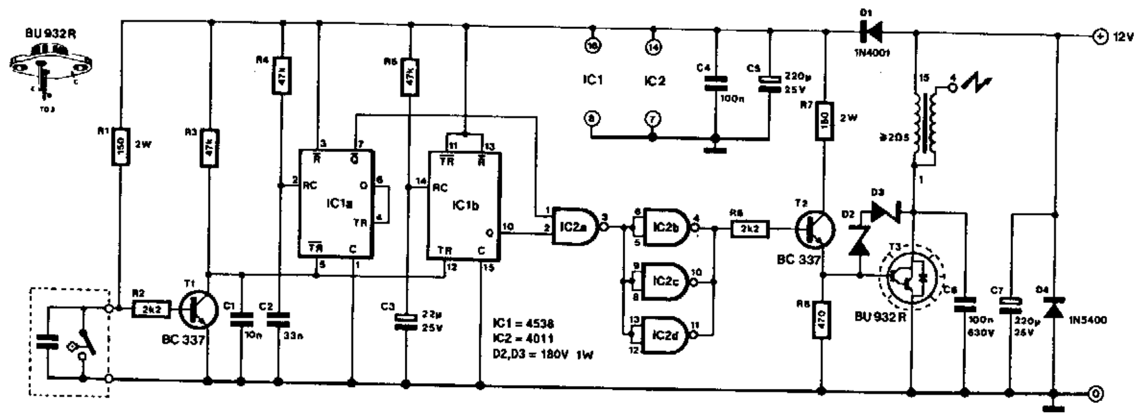
Fig. 12-1

In the wiper-delay schematic shown, with the ignition on, D1 maintains regulated +12 Vdc. When S1 closes, C1 bypasses transients and passes this +12 Vdc to divider R2-R3, producing a TTL high at pin 3 of IC1, a 4013 CMOS dual leading-edge triggered D-type flip-flop. Filter R4/C3 keeps IC1 from triggering erroneously when the ignition is on. When S1 is pressed, output Q1 (pin 1) of IC1 latches high, turning on Q1, which conducts via R5, turning on IC2; LED1 indicates power, and R6 sets the current. Because IC2 depends on Q1 for power, IC2 stays off until Q1 turns on.

The left half of IC2 is an astable, with its delay set by R7, R8, R11, and C4. The right half of IC2 is a monostable, with its pulse duration set by R9 and C5. With the values used, you might expect R11 to vary the delay from about 15 to 84 seconds, with a 2.42-second monostable pulse operating the wiper blades on each cycle. However, the actual delay will range between 2 to 18 s, with a 1-s monostable pulse on each cycle. That discrepancy stems from the fact that IC2 is being fed from the emitter of Q1, rather than directly from the regulated +12-V supply. Transistor Q1 acts as an active current source, charges and discharges C4 faster than it ordinarily would.

The astable output (OUT1, pin 5) is tied to TRIG2 (pin 8). When OUT2 (pin 9) becomes high, Q2 is biased via R10 and current flows through RY1, with D2 dissipating back-emf when RY1 shuts off.

ELECTRONIC IGNITION



ELEKTOR ELECTRONICS

Fig. 12-2

This electronic ignition circuit is intended to be inserted into a car's conventional ignition system. In effect, it replaces the original 12-V switching circuit in the primary winding of the coil by one generating more than 100 V. It thereby converts a current circuit, which is upset by lead and stray resistance, into a voltage circuit that is much more efficient.

The pulses emanating from the contact breaker, shown at the extreme lower left-hand side of the diagram, are applied to transistor T1 and subsequently differentiated by R3/C1. This causes a negligible ignition delay. The current through the contact-breaker points is determined by the value of R1. This value has been chosen to ensure that the points remain clean.

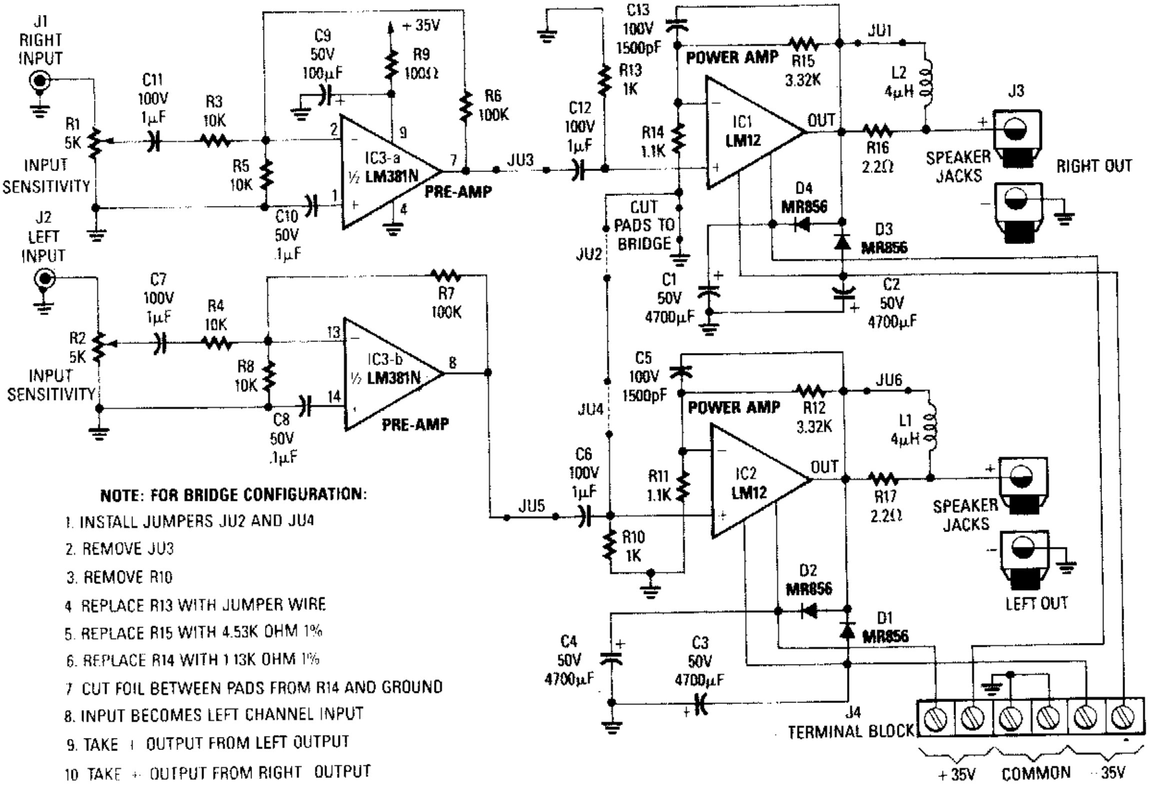
Transistor T1 is followed by two monostables, IC1A and IC1B, which are both triggered by the output pulses of T1. However, whereas IC1A is triggered by the trailing edge, IC1B is triggered by the leading edge.

Monostable IC1A passes a pulse of about 1.5 ms (determined by R4/C2) to NAND gate IC2A. This gate switches off high-voltage Darlington T3 via gates IC2B, IC2C and IC2D, and driver T2, for the duration of the pulse. Gate IC2 ensures that T3 is switched on only when the engine is running, to prevent a current of some amperes flowing through the ignition coil.

As long as pulses emanate from the contact breaker, IC1B is triggered and its *Q* output remains logic high. The mono time of this stage is about 1 s and is determined by R5/C3.

Darlington T3 is switched on via T2 and IC2A through IC2D as long as IC1A does not pass an ignition pulse. When the engine is not running, the *Q* output of IC2B goes low after 1 s and this causes T2 and T3 to be switched off. The two series-connected 180-V zener diodes protect the collector of the BU932R against too high of a voltage. The Darlington must be fitted on a suitable heatsink.

HIGH-POWER CAR AUDIO AMPLIFIER

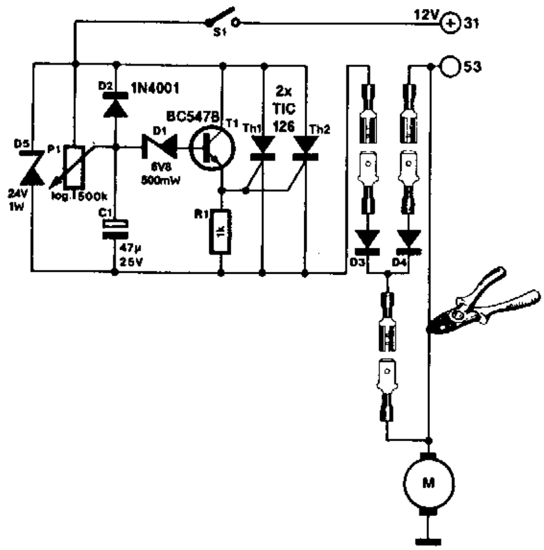


RADIO-ELECTRONICS

Fig. 12-3

This stereo amp will supply 60 W rms into 8 Ω or 100 W rms into 4 Ω. Notice that the LM12C1 line level (about 300 mV) into 5 KΩ. A ±35-V supply is required for full power output. Power can be obtained from a dc-dc converter.

WINDSHIELD WIPER INTERVAL CONTROLLER



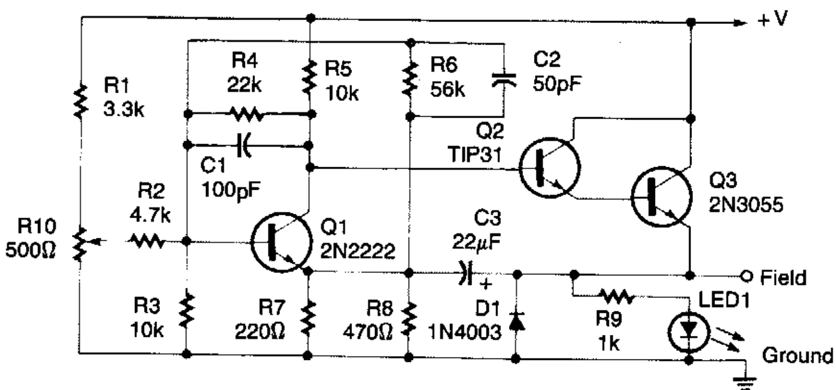
ELEKTOR ELECTRONICS *Fig. 12-4*

The windshield wiper interval circuit presented here is very compact and is noteworthy for its use of two thyristors, instead of a relay. It has only two connections and operates without any problems—even in conjunction with multistage wiper circuits.

The connecting wire between the wiper motor and terminal 53 is cut and new connections are made (as shown in the diagram). When the interval switch, S1, is closed, capacitor C1 charges via P1 and the wiper motor. After a time set with P1, transistor T1 switches on and triggers the thyristors. The wiper motor is then energized via the thyristors and D3 and sets the wipers into motion. At the same time, C1 discharges via D2 and the thyristors.

After a short time, the wiper stop switch connects terminal 53 to the +12-V line so that the wiper motor is energized via D4. The thyristors are switched off because the voltage drop across D3 plus Th1/Th2 is then greater than that across D4. When the wipers reach the end of their travel again, the stop switch connects terminal 53 to ground and enables C1 to charge again.

VOLTAGE REGULATOR FOR CARS AND MOTORCYCLES



RADIO-ELECTRONICS **Fig. 12-5**

This regulator circuit can be used on an alternator that has one field terminal grounded. When +V (input) gets too high, Q1 conducts, and the base of Q2 is driven toward ground, reducing the voltage fed to Q3. This lowers the voltage fed to the field of the alternator.

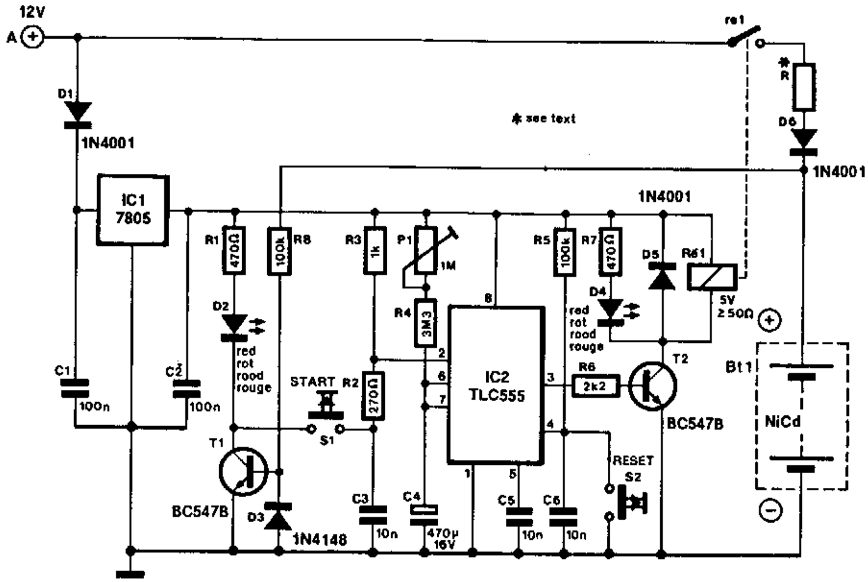
13

Battery Chargers

The sources of the following circuits are contained in the Sources section, which begins on page 662. The figure number in the box of each circuit correlates to the entry in the Sources section.

Portable Nicad Charger
Battery Charger
Current-Limited 6-V Charger
12-V Battery Charger
+12-Vdc Mobile Battery Charger
Charger Extends Lead-Acid Battery Life

PORTABLE NICAD CHARGER



ELEKTOR ELECTRONICS

Fig. 13-1

The portable charger is intended primarily to give model enthusiasts the opportunity of charging their Nicad batteries from a car battery out in the open. The supply voltage for the circuit is regulated by IC1.

When the circuit is connected to the car battery, D2 lights only if the Nicad to be charged has been connected with correct polarity. For that purpose, the + terminal of the Nicad battery is connected to the base of T1 via R8. Because even a discharged battery provides some voltage, T1 is switched on and D2 lights.

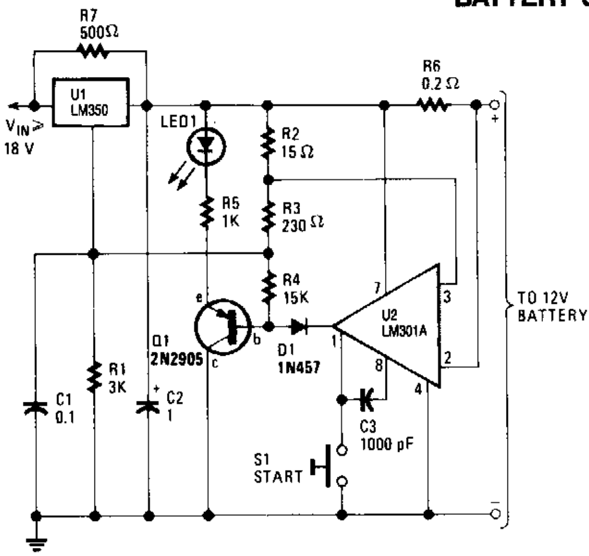
Only if the polarity is correct will the pressing of the start switch, S1, have any effect. If so, the collector voltage of T1 is virtually zero so that monostable IC2 is triggered by S1. The output, pin 3, of this CMOS timer then becomes high, T2 is switched on and relay Re1 is energized. Charging of the Nicad battery, via R5 and D6, then begins and charging indicator D4 lights. During the charging, C4 is charged slowly via P1 and R4. The value of these components determines the mono time of IC2 and thus the charging period of the Nicad battery. With values as shown in the diagram, that period can be set with P1 to between 26 and 33 min. Notice that this time is affected by the leakage current of C4; use a good-quality capacitor here. The charging can be interrupted with reset switch S2.

The charging current through the Nicad battery is determined by the value of R, which can be calculated:

$$R = \left\{ \frac{12 - (0.7 + 1.3x\,\text{no. of cells})}{I_c} \right\} [\Omega]$$

I_c is the charging current, which is here because the chosen charging period is twice the nominal value of the capacity of the Nicad battery. Resistor R must be able to dissipate a power of $I_c^2\,R$ W. Finally, make sure that the Nicad battery is suitable for fast charging; never charge for longer than half an hour!

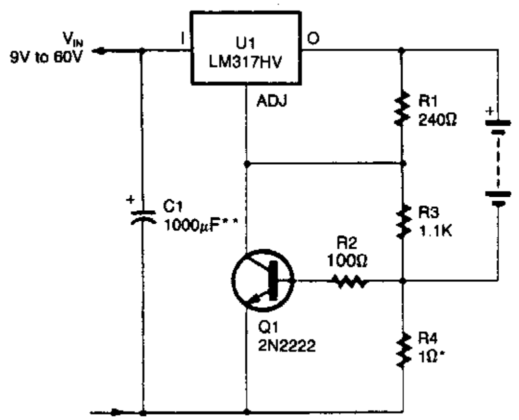
BATTERY CHARGER



This high-performance charger quickly charges gelled lead-acid batteries, and turns off at full charge. At first, the charge current is held at 2 A, but as battery voltage rises, current decreases. When current falls to 150 mA, the charger automatically switches to a lower float voltage to keep from overcharging. When you hit full charge, transistor Q1 lights the LED to indicate that status.

HANDS-ON ELECTRONICS *Fig. 13-2*

CURRENT-LIMITED 6-V CHARGER

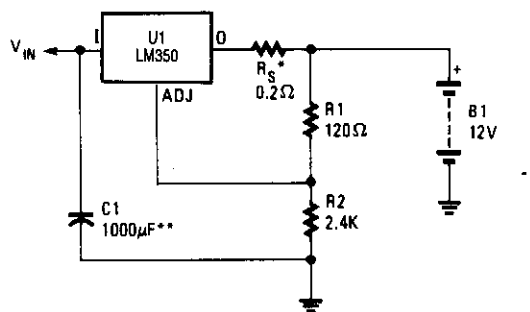


*Sets Peak Current (0.6A for 1Ω)
**The 1000µF is Recommended to Filter Out Input Transients.

POPULAR ELECTRONICS *Fig. 13-3*

An LM317HV regulator is used as a current-limited charger. If current through R4 exceeds 0.6 A, Q1 is biased on, which pulls the ADJ terminal of the LM317 HV to ground and reduces the battery-charging current.

12-V BATTERY CHARGER

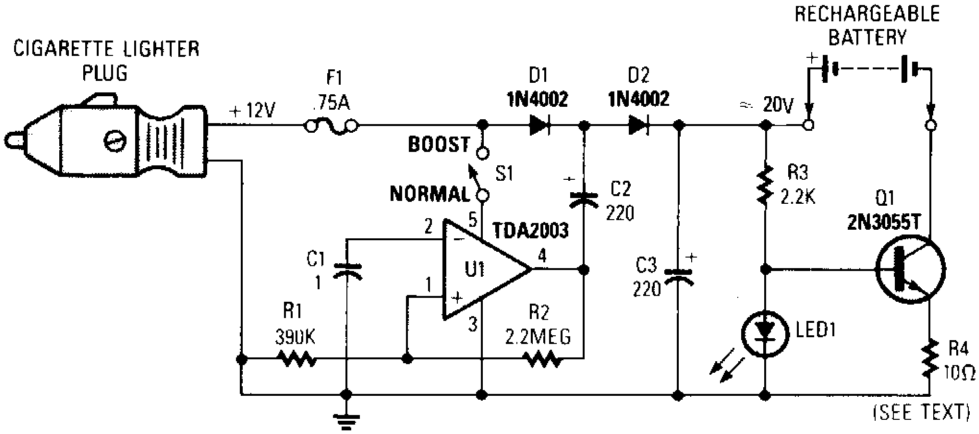


*R_S–SETS OUTPUT IMPEDANCE OF CHARGER $Z_{OUT} = R_S\,(1 + \frac{R2}{R1})$
USE OF R_S ALLOWS LOW CHARGING RATES WITH FULLY CHARGED BATTERY.
**THE 1000µF IS RECOMMENDED TO FILTER OUT INPUT TRANSIENTS.

POPULAR ELECTRONICS *Fig. 13-4*

This simple charger uses an LM350 regulator as a battery charger.

+ 12-Vdc MOBILE BATTERY CHARGER

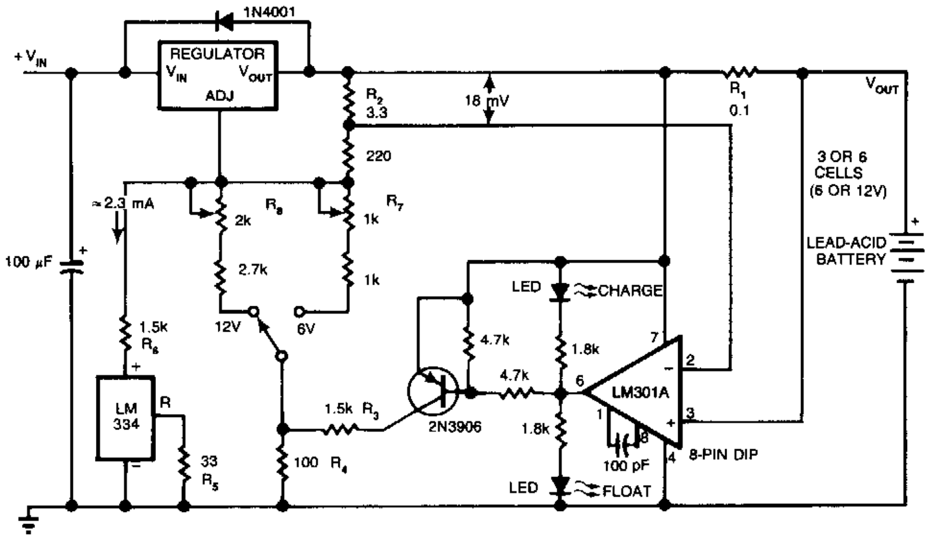


Cell Size	Amp/hr Rate	R4 value (14 hr rate)
N	150 mA	120 ohms @ .25-watt
AA	500 mA	47 ohms @ .5 watt
C	1500 mA	12 ohms @ .5 watt
D	1500 mA	12 ohms @ .5 watt
D (High capacity)	4000 mA	3.3 ohms @ 2 watt

POPULAR ELECTRONICS *Fig. 13-5*

This circuit provides up to 20 V output from a 12-V automotive supply, to enable constant current charging of Nicad battery assemblies up to about 18 V total. V1 forms a square-wave oscillator, D1 and D2, coupling this square wave to the 12-V battery supply to obtain over 20 Vdc. If this is not needed, S1 is left open. Q1 forms a current regulator to determine the charging rate of the rechargeable battery. R4 is selected from the table or it can be switched with a rotary selector switch.

CHARGER EXTENDS LEAD-ACID BATTERY LIFE



EDN

Fig. 13-6

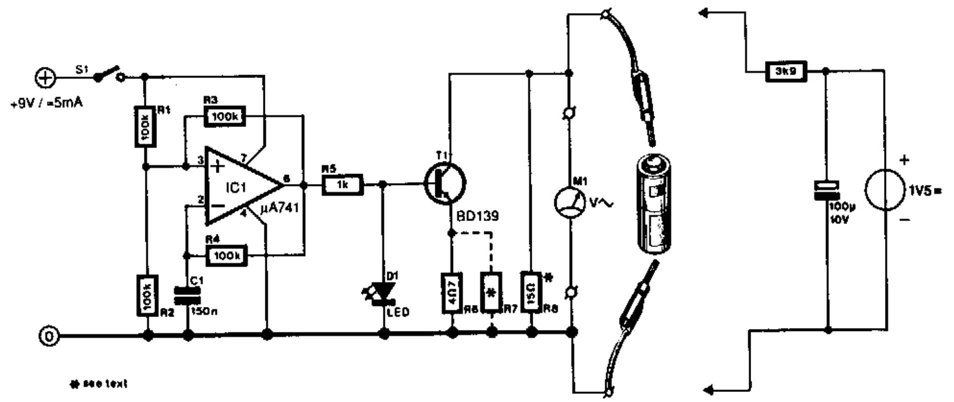
The circuit furnishes an initial charging voltage of 2.5 V per cell at 25°C to rapidly charge a battery. The charging current decreases as the battery charges, and when the current drops to 180 mA, the charging circuit reduces the output voltage to 2.35 V per cell, floating the battery in a fully charged state. This lower voltage prevents the battery from overcharging, which would shorten its life. The LM301A compares the voltage drop across R1 with an 18-mV reference set by R2. The comparator's output controls the voltage regulator, forcing it to produce the lower float voltage when the battery-charging current passing through R1 drops below 180 mA. the 150-mV difference between the charge and float voltages is set by the ratio of R3 to R4. The LEDs show the state of the circuit.

14

Battery Monitors

The sources of the following circuits are contained in the Sources section, which begins on page 662. The figure number in the box of each circuit correlates to the entry in the Sources section.

INTERNAL RESISTANCE BATTERY TESTER



ELEKTOR ELECTRONICS

Fig. 14-1

A designer often needs to know the value of the internal resistance of a battery. Quite a few testers give a relative indication of the value, but this is seldom in ohms. The present tester can, in principle, provide that information.

The basic idea behind it is to load the battery with a varying current, so as to cause an alternating-voltage drop across the internal resistance that can be measured at the battery terminals. Provided that current variations are regular and constant, the voltage drop is directly proportional to the internal resistance.

Choose the variation of the current carefully to read the value of the internal resistance directly on the scale of an ac voltmeter.

The load current is varied with the aid of a current source, T1 in the diagram, which is switched on and off by square-wave generator IC1. The chosen switching frequency of 50 Hz ensures that the ac component at the battery terminals can be measured by a standard ac voltmeter (universal meter).

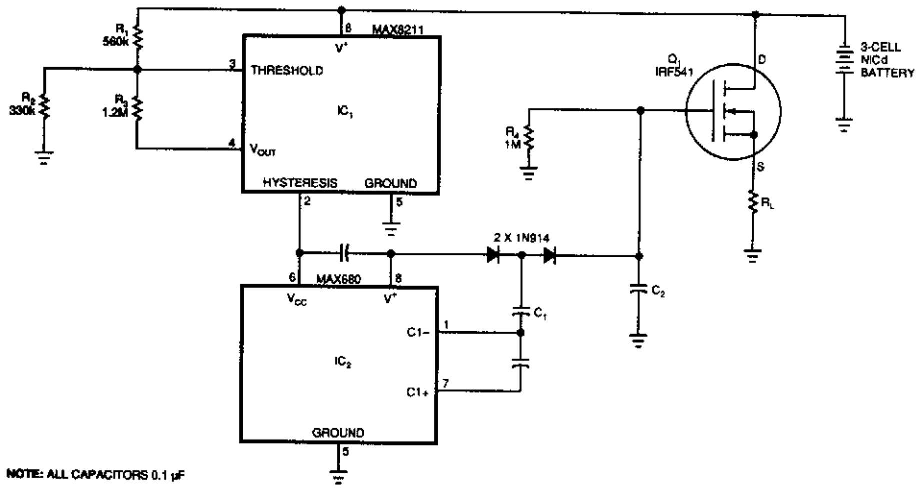
The battery is loaded constantly by R8, which has a value of 1.5 Ω for 1.5-V batteries, shunted by the ac voltmeter. The indicated voltage times 10 is the value of the internal resistance of the battery. When the battery under test is flat or if the supply battery is flat, no current flows and the meter will read zero. It would then appear as if the battery under test is an ideal type—without internal resistance.

A flat supply battery is indicated if D1 does not light. You can ascertain that the battery under test is flat by measuring the direct voltage across its terminals. The load must be left connected, of course, otherwise the emf is measured and this may well be 1.5 V—even if the battery is flat.

The tester is calibrated with the aid of the auxiliary circuit (shown at the extreme right in the circuit diagram). The 1.5-V supply and electrolytic capacitor form a virtually ideal voltage source, of which the 3.9-Ω resistor forms the internal resistance. With this source connected across the output terminals of the tester, a suitable value should be ascertained for R7. That value is found when the ac voltmeter shows 0.39 V. Notice that this procedure is not the same for all measuring instruments: the alternate use of the digital and a moving coil meter, for instance, is not feasible.

The tester is intended for 1.5-V batteries. The load current is fairly high: about 100 mA through R8 and around 170 mA through T1. For 9-V batteries that current is too high: the current should then be reduced by taking greater values for R6 through R8.

BATTERY-SAVING DISCONNECT SWITCH



NOTE: ALL CAPACITORS 0.1 µF

EDN

Fig. 14-2

At a predetermined level of declining terminal voltage, the circuit disconnects the battery from the load and halts potentially destructive battery discharge. Q1, a high-side, floating-source MOSFET, acts as the switch. The overall circuit draws about 500 μA when the switch is closed and about 8 μA when it's open.

The values of R_1, R_2, and R_3 set the upper and lower voltage thresholds, V_U and V_L, according to the relationships

$$R_1 = R_2\left(\frac{V_L}{1.15}\right) - 1$$

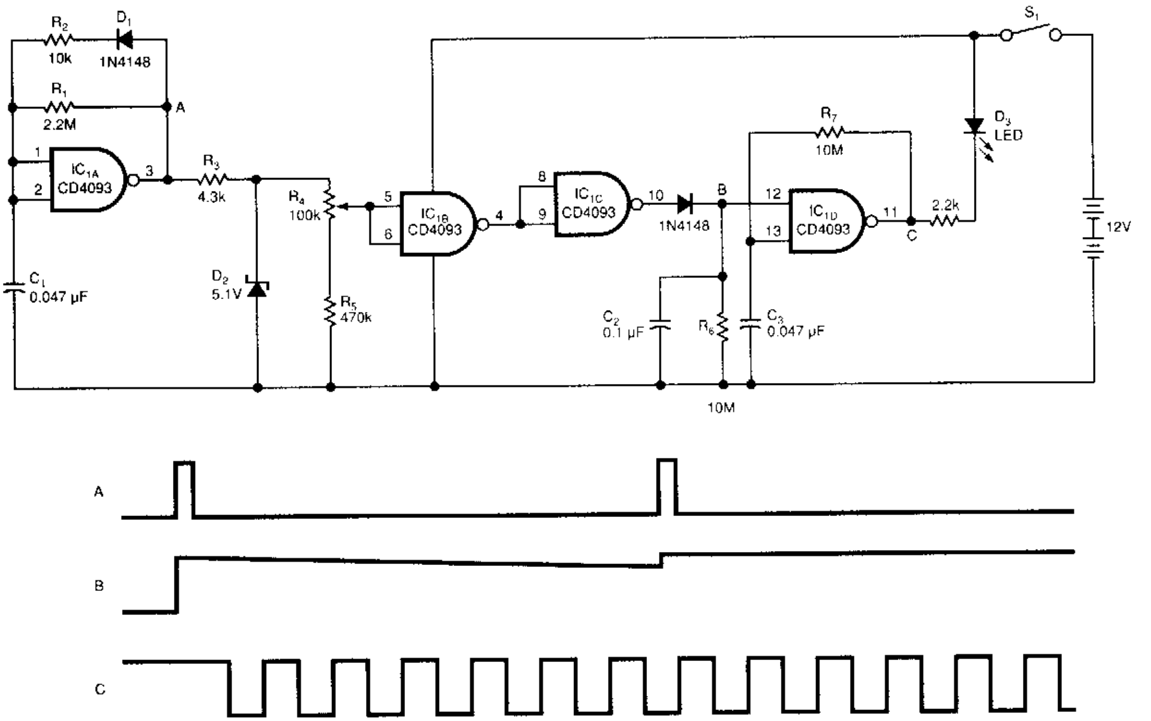
and

$$R_3 = 1.15\left(\frac{R_1}{V_U - V_L}\right)$$

For the circuit to start, $V+$ must exceed V_U. The voltage detector IC1 then powers IC2, but only while $V+$ remains above V_L. Otherwise, IC2 loses its power, removes gate drive from Q1, and turns it off. IC2 is a dual charge-pump inverter that normally converts 5 V to $\pm$10 V. Capacitors C1, C2, and the two associated diodes form a voltage tripler that generates a gate drive for Q1 that is approximately equal to two times the battery voltage.

With the values in the schematic, the circuit disconnects 3-cell Nicad battery from its load when $V+$ reaches a V_L of 3.1 V. Approximately 0.5 V of hysteresis prevents the switch from turning on immediately when the circuit removes the load; $V+$ must first return to V_U, which is 3.6 V. The gate drive declines as the battery voltage declines, cause the ON-resistance of Q1 to reach a maximum of approximately 0.1 Ω, just before $V+$ reaches its 3.1-V threshold. A 300-mA load current at that time will cause a 30-mV drop across the disconnect switch. The drop will be 2 to 3 mV less for higher battery voltages. Resistor R4 ensures that Q1 can adequately turn off by providing a discharge path for C2.

LOW-BATTERY DETECTOR



EDN

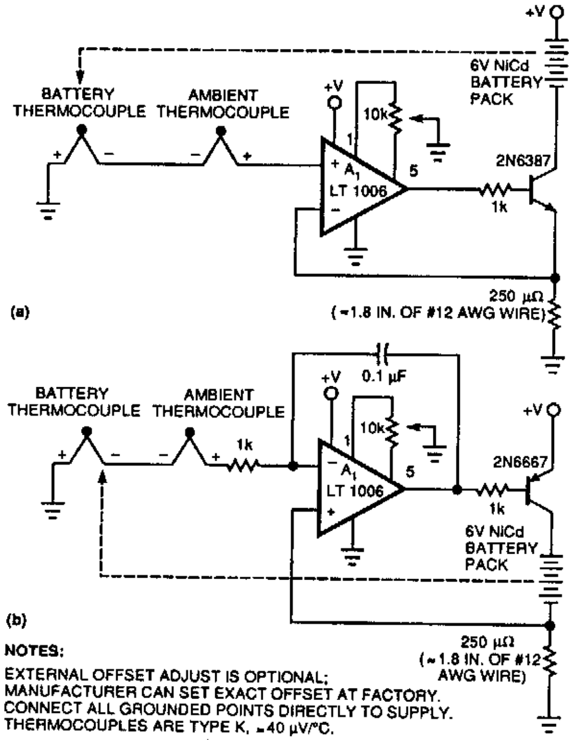
Fig. 14-3

The battery low-voltage detector uses a CD4093 Schmitt trigger and a capacitor that acts as a 1-bit dynamic RAM. The circuit conserves power by using a periodic test method. IC1A, C1, R1, R2, and D1 generate a narrow, positive pulse at point A.

D2, R4, and R5 regulate and divide the signal at A. Thus, the input of IC1B is independent of the power supply. Because the threshold voltage of the Schmitt trigger depends on the power supply, the threshold voltage will drop if the power-supply voltage drops. When the threshold voltage is lower than the input voltage, IC1B will become low, and IC1C's output will become high.

Capacitor C2 stores the results of the periodic test. The time constant C2 and R6 set is 1 s, and the test period is approximately 0.1 s. When point B is high, which implies that the battery is low, IC1D, C3, and R7 generate a square waveform, which lights D3. You can adjust the detected voltage level by adjusting R4. You can test different battery voltages by changing the voltage level of D2.

BATTERY-TEMPERATURE SENSING NICAD CHARGER

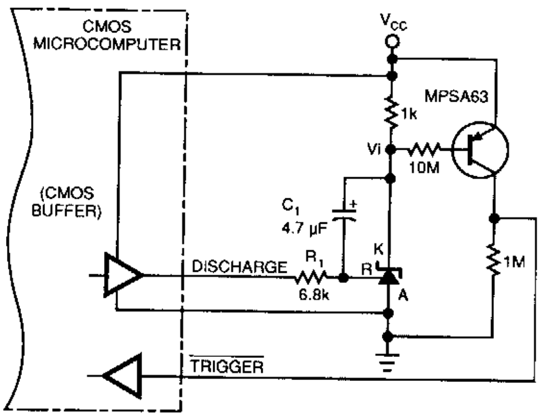


EDN

Fig. 14-4

Two simple circuits permit Nicad charging of a battery based on temperature differences between the battery pack and the ambient temperature. This method has the advantage of allowing fast charging because the circuit senses the temperature rise that occurs after charging is complete and the battery under charge is producing heat, not accumulating charge.

BATTERY-VOLTAGE MEASURING REGULATOR

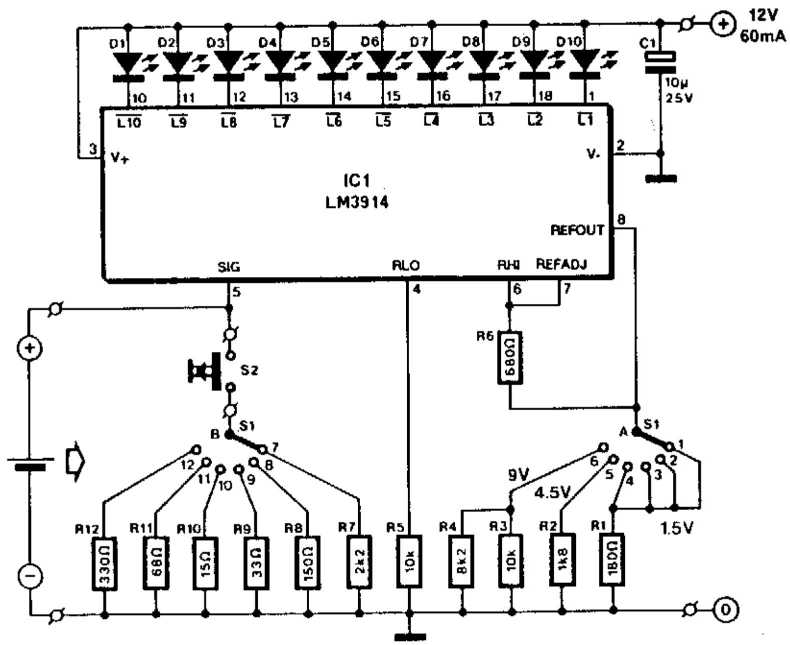


EDN

Fig. 14-5

This circuit allows a microprocessor system to measure its own battery voltage. A Texas Instrument TI431 precision shunt regulator acts as a precision reference and integrator/amplifier, measuring its own supply via voltage-dependent charge/discharge time intervals. Notice that you must write a short control and voltage calculation software routine for your system.

BATTERY TESTER



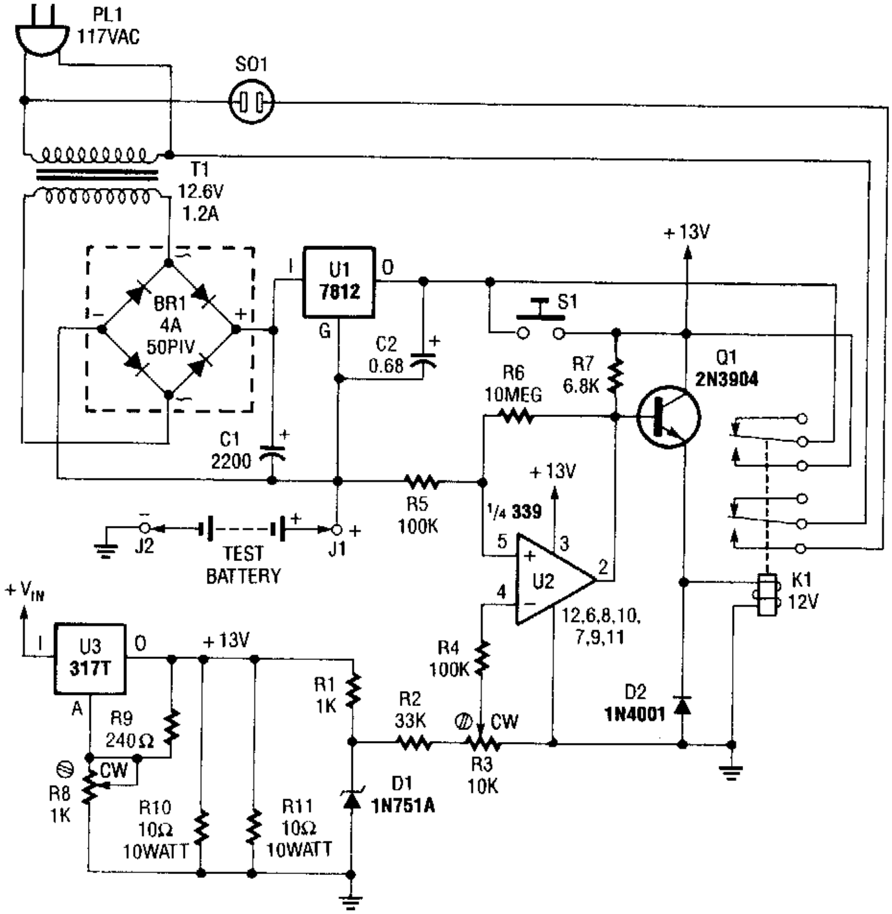
Range	LED									
	D1	D2	D3	D4	D5	D6	D7	D8	D9	D10
	red	orange		green						
1.5 V	0.86	0.96	1.04	1.13	1.21	1.29	1.38	1.46	1.55	1.63
4.5 V	2.58	2.83	3.05	3.31	3.57	3.82	4.07	4.33	4.57	4.82
9.0 V	5.3	5.8	6.3	6.9	7.4	7.9	8.5	9.0	9.5	10.2

ELEKTOR ELECTRONICS

Fig. 14-6

This battery tester makes use of an LM3914 bar-graph driver IC. S1 selects load on battery under test and programs the voltage range. S2 loads the battery under test. The table gives the calibration factors for the tester. LEDs D1 through D10 are used as indicators.

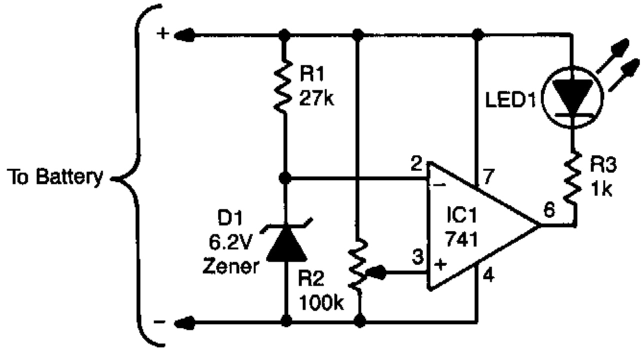
NICAD BATTERY TESTER



POPULAR ELECTRONICS *Fig. 14-7*

This Nicad battery tester discharges the test battery at a rate of 500 mA. When the endpoint of 1 V (determined by setting of R3) is resolved, pin 2 of U2 becomes low, deactivating Q1 and disconnecting the test battery from the circuit. Power for U3 comes from the 12-V regulator in series with the battery being tested. A clock or timer can be plugged into S1 to indicate the time it takes to discharge the battery under test.

LOW-BATTERY INDICATOR

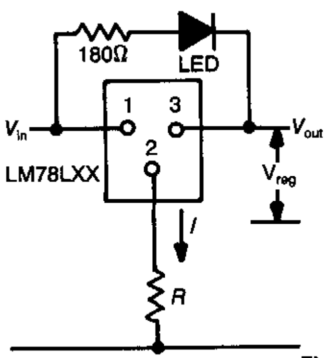


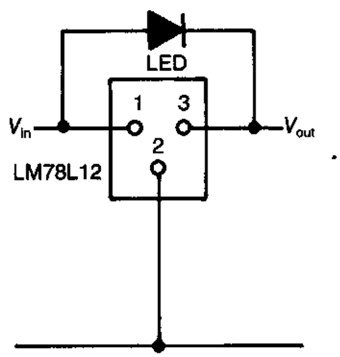
42 NEW IDEAS

Fig. 14-8

The sensing circuit consists of a 741 op amp set up as a voltage comparator, using a zener diode as a voltage reference. The op amp is inserted as a bridge between two resistance ladders, one which contains the zener reference, and the other a high-value linear potentiometer. When the voltage at the wiper of the potentiometer drops below the voltage set by the zener, the output of the op amp becomes low; that turns on the LED connected between it and V_{CC}. The circuit can be adapted to work with battery-powered circuits that require between 6 and 18 V; the only changes needed would be a lower-voltage zener and a smaller current-limiting resistor in the case of voltage below 9 V, and a larger resistor for higher voltages.

VOLTAGE-LEVEL INDICATOR





ELECTRONIC ENGINEERING

Fig. 14-9(a)

Fig. 14-9(b)

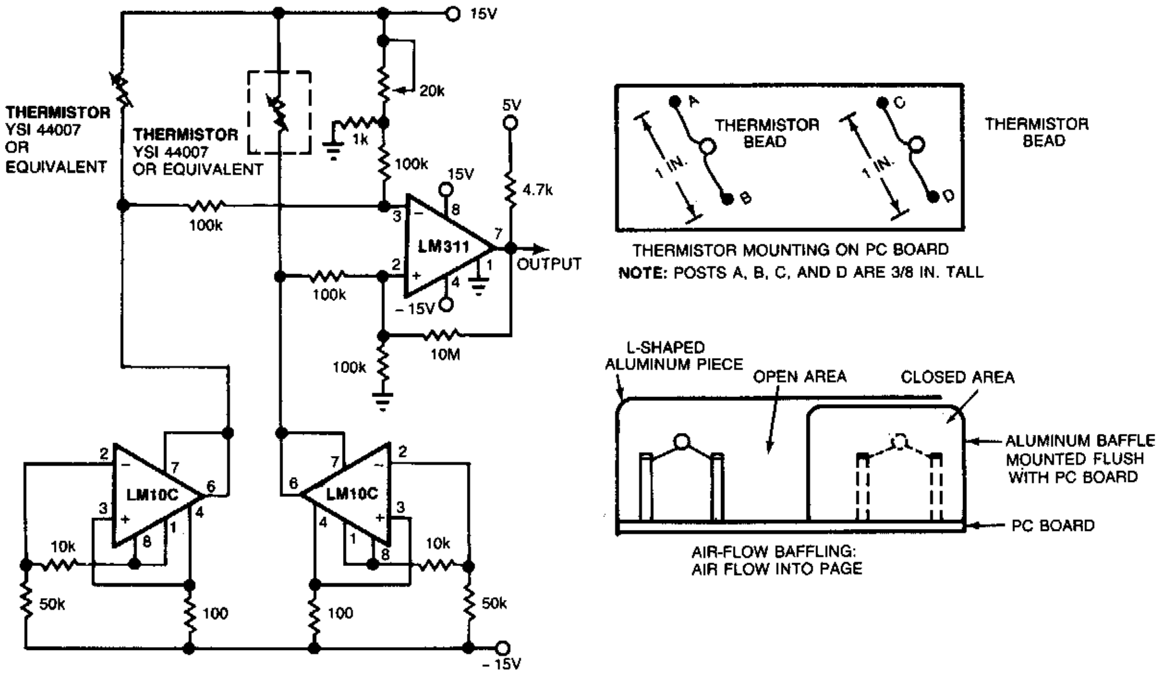
Three-terminal regulator device (LM78LXX) has $V_{out} = V_{in}$ until the input rises 1.5 to 2 V above the output when the regulated voltage $V_{reg} = XX$ is obtained. A differential of 1.5 V between input and output is necessary to light the LED. Thus, the LED lights when V_{in} rises above $V_{reg} + IR + 1.5$ V, where I is typically 6 mA (a zener diode could be used in place of R). For input voltages much higher than necessary to light the LED, a current-limiting resistor in series might be necessary. A useful automotive application is shown in Fig. 14-9(b). The circuit indicates when battery voltage is above 13.5 V which indicates (in conjunction with an ammeter) whether the alternator/regulator/battery system is operating correctly. With the engine off, the battery voltage drops to 12 V and the LED extinguishes. The circuit requires no calibration.

15

Bridge Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 663. The figure number in the box of each circuit correlates to the entry in the Sources section.

AIR-FLOW SENSING THERMISTOR BRIDGE



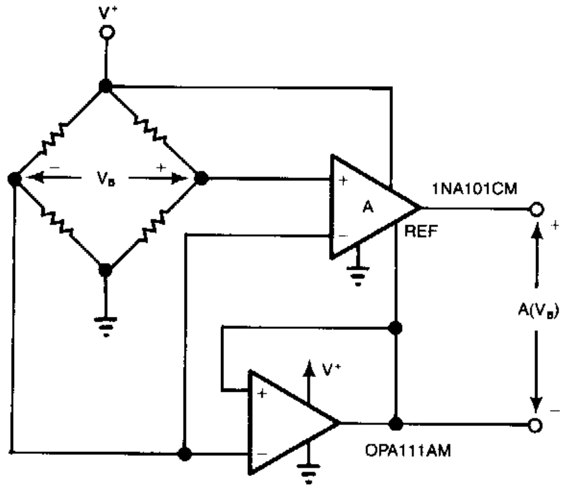
EDN

Fig. 15-1

Using the thermistor-bridge circuit, you can detect system-cooling air losses caused by filter or inlet blockage or fan failure. One thermistor is mounted directly in the air flow; the other is baffled. The exposed thermistor senses the temperature in the cooling system; the baffled thermistor senses the ambient temperature in still air. As long as the thermistors are at different temperatures, the bridge stays unbalanced and the circuit produces a logical high, indicating that the cooling system is working. If the air flow stops, the exposed thermistor will reach ambient temperature, the bridge will become balanced, and the circuit will indicate ventilation-system failure by producing a logical low.

The bridge circuit's matched thermistors are biased by matched-current sources. Two LM10C operational amplifiers act as constant-current sources, and an LM311 comparator senses the difference between the voltage drops across the thermistors, producing the logical high when the bridge is unbalanced and the logical low when the bridge is balanced. Use a 20-kΩ potentiometer to set the comparator's threshold; this setting determines the minimum air flow that will cause the circuit to produce a logical high.

BRIDGE CIRCUIT WITH ONE POWER SUPPLY

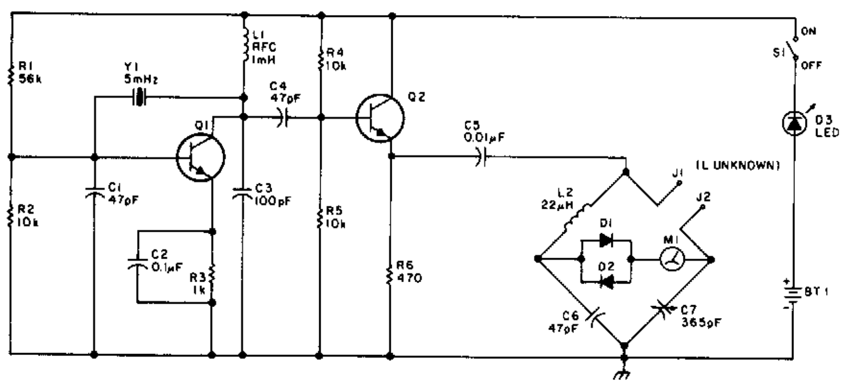


EDN *Fig. 15-2*

For systems with only one power supply, two op amps act as instrumentation and buffer amps. The OPA111AM buffers the reference mode of the bridge and applies that voltage to the instrumentation amps REF terminal. Output is taken between the amplifier outputs to exclude the fixed output offset.

The additional op amp creates a bridge error of $I_B \times R/2$, where I_B=bias current of op amp and R is the resistance of one leg of the bridge.

INDUCTANCE BRIDGE



73 AMATEUR RADIO *Fig. 15-3*

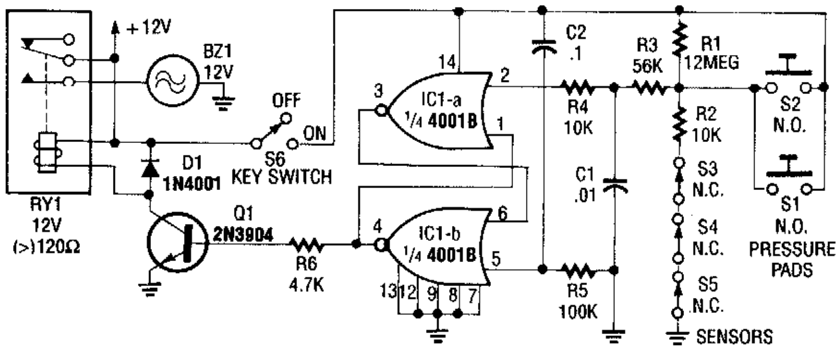
This bridge will measure inductances from about 1 to 30 μH at a test frequency of 5 mHz. A 365-pF AM-type tuning capacitor is used as a variable element. The circuit should be constructed in a metal enclosure. Calibration can be done on known inductors or by plotting a curve of the capacitance of the 365-pF capacitor versus rotation and calculating the inductance from this. The range of measurement can be charged by using a different frequency crystal and/or variation of L2 and C6.

16

Burglar Alarms

The sources of the following circuits are contained in the Sources section, which begins on page 663. The figure number in the box of each circuit correlates to the entry in the Sources section.

SELF-LATCHING BURGLAR ALARM



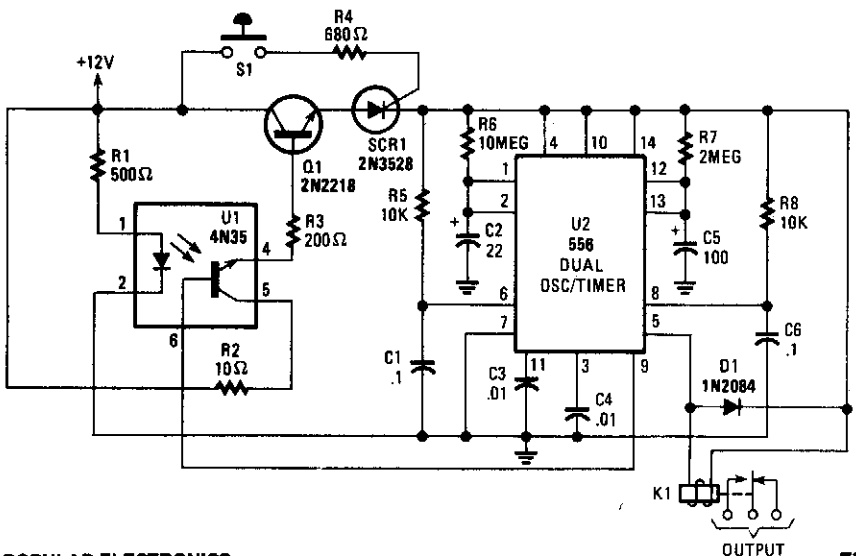
A SIMPLE SELF-LATCHING BURGLAR ALARM.

RADIO-ELECTRONICS

Fig. 16-1

This alarm uses IC1A and IC1B as a latch. When sensors S1 through S5 activate, IC1A turns on and forces IC1B to cut off. Q1 drives RY1.

BURGLAR ALARM WITH TIMED SHUTOFF



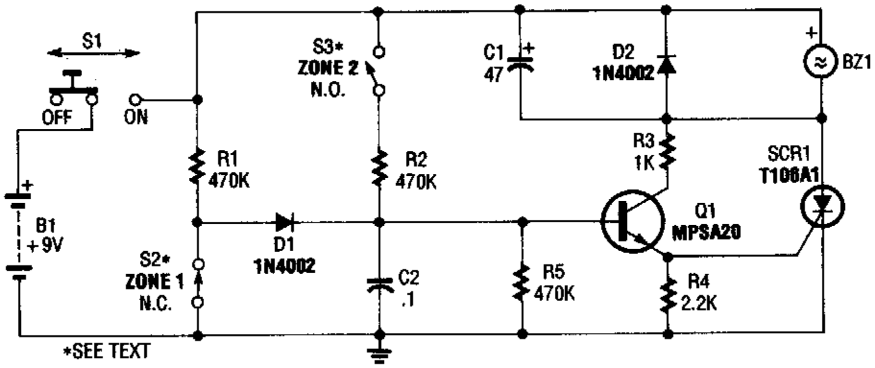
POPULAR ELECTRONICS

Fig. 16-2

When S1 (sensor) is closed, power is applied to U2, a dual timer. After a time determined by C2, C1 is energized after a predetermined time determined by the value of C5, pin 9 of U2 becomes low, switching off the transistor in the optoisolater, cutting anode current of SCR1 and de-energizing K1. The system is now reset. Notice that $(R_6 \times C_2)$ is less than $(R_7 \times C_5)$. The ON time is approximately given by:

$$(R_7 \times C_5) - (R_6 \times C_2) = t_{ON}$$

SIMPLE BURGLAR ALARM

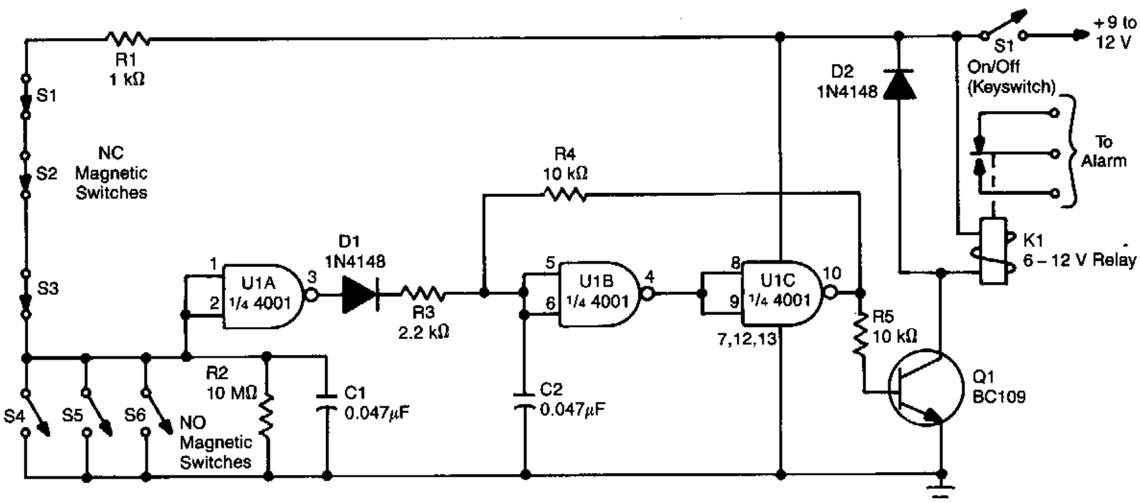


POPULAR ELECTRONICS

Fig. 16-3

A simple circuit using either NO or NC sensors uses an RC delay circuit (R2/C2 or R1/C2) to drive emitter-follower Q1, switching SCR1 and buzzer (or bell) BZ1. S1 is used for activation and reset.

SIMPLE BURGLAR ALARM

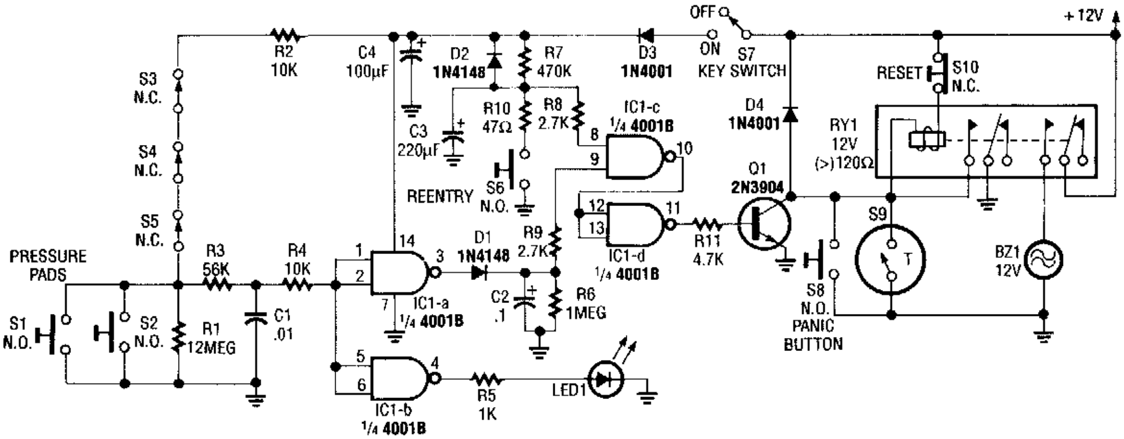


POPULAR ELECTRONICS

Fig. 16-4

Using one IC and a driver transistor, this simple alarm uses either NO or NC sensors. When a sensor operates, the input to U1A goes low, causing U1A to go high, U1B low, and U1C high. This biases Q1 ON and activates relay K1. On/off is via keyswitch S1.

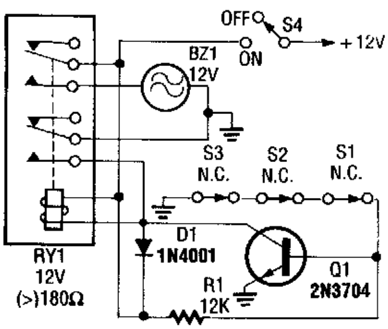
HOME SECURITY SYSTEM



RADIO-ELECTRONICS *Fig. 16-5*

This alarm circuit activates when S1 through S5 are activated. This lights LED1 and activates Q1 via IC1C and IC1D. RY1 is wired to self latch. S10 is used to reset. When key switch S1 is activated or when re-entry buttons at S6 are depressed, IC1C is deactivated until RC network R7/C3 charges.

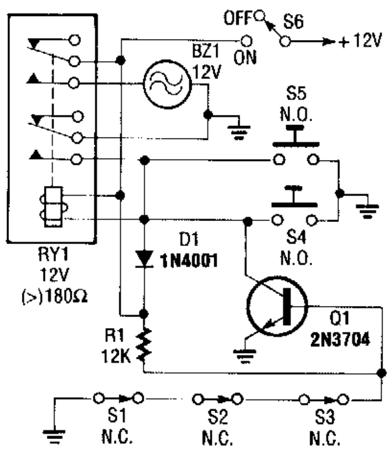
SIMPLE BURGLAR ALARM WITH NC SWITCHES



RADIO-ELECTRONICS *Fig. 16-6*

This relay draws 1 mA of idling current. Q1 detects open switch and energizes RY1.

BURGLAR ALARM WITH NC AND NO SWITCHES



RADIO-ELECTRONICS *Fig. 16-7*

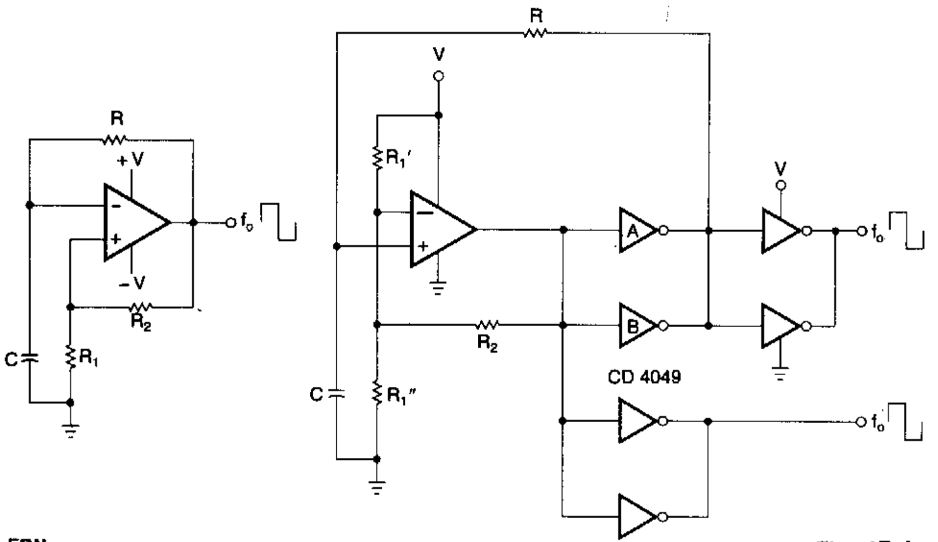
This circuit uses both NC and NO sensors. Series NC sensors allow Q1 to activate RY1. NO sensors directly activate RY1.

17

Buffers

The sources of the following circuits are contained in the Sources section, which begins on page 663. The figure number in the box of each circuit correlates to the entry in the Sources section.

OSCILLATOR BUFFERS

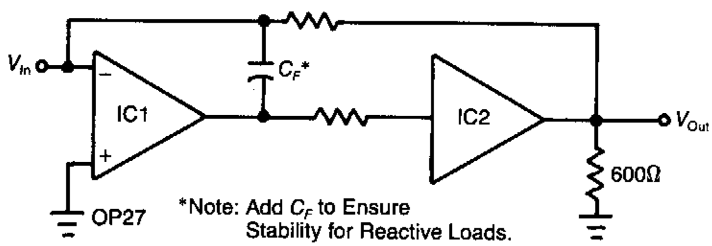


EDN

Fig. 17-1

CMOS buffers added to an op amp oscillator improve performance, largely as a result of nonsymmetry and variability of the op amp's output saturation voltages.

PRECISION-INCREASING BUFFER



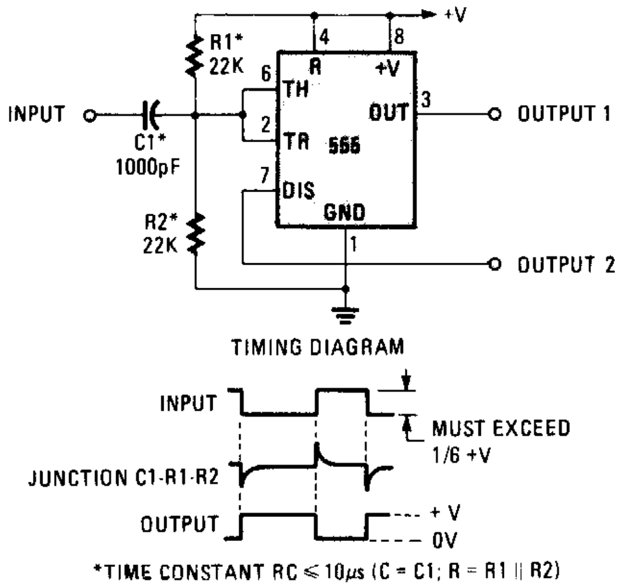
EDN

Fig. 17-2

Adding an unity-gain buffer to your analog circuit can increase its precision. For example, by itself, the op amp IC1 exhibits a maximum dV_{OS}/dT of 1.8 $\mu V/°C$ and can drive a 600-Ω load. Under these conditions, IC1 would dissipate 94 mW incrementally. Thus, the op amp's O_{JA} of 150°C/W would change its V_{OS} by 25 μV.

The buffer, IC2, will isolate IC1 from the load and eliminate the change in power dissipation in IC1, thereby achieving IC1's minimum, rated offset-voltage drive. The loop gain of IC1 essentially eliminates the offset of the buffer. Almost any unity-gain buffer will work, provided that it exhibits a 3-dB bandwidth that is at least 5 times the gain-bandwidth product of the op amp.

INVERTING BISTABLE BUFFER



HANDS-ON ELECTRONICS *Fig. 17-3*

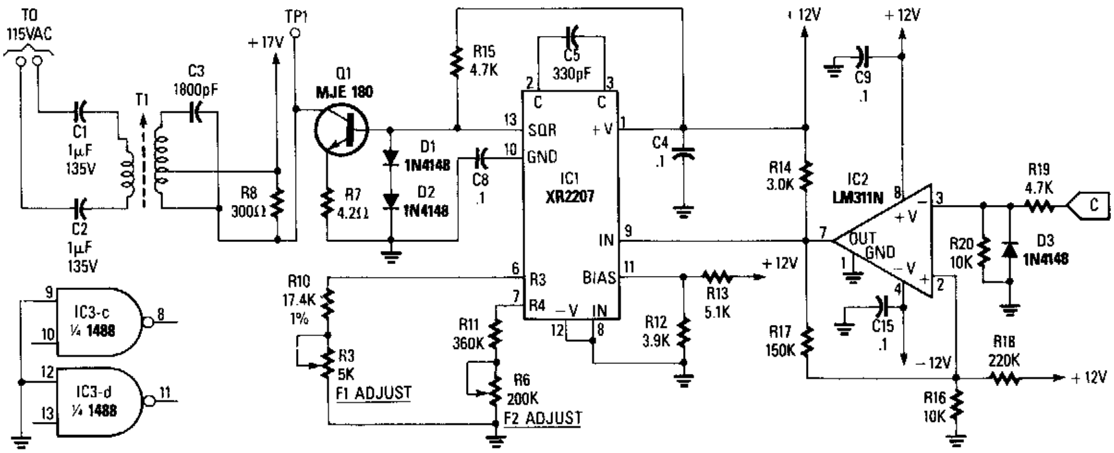
This circuit uses a 555 timer as a flip-flop bistable buffer.

18

Carrier-Current Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 663. The figure number in the box of each circuit correlates to the entry in the Sources section.

CARRIER CURRENT TRANSMITTER FOR DATA TRANSMISSION



RADIO-ELECTRONICS

Fig. 18-1

$$f_1 = \frac{1}{R_3 C_1}$$

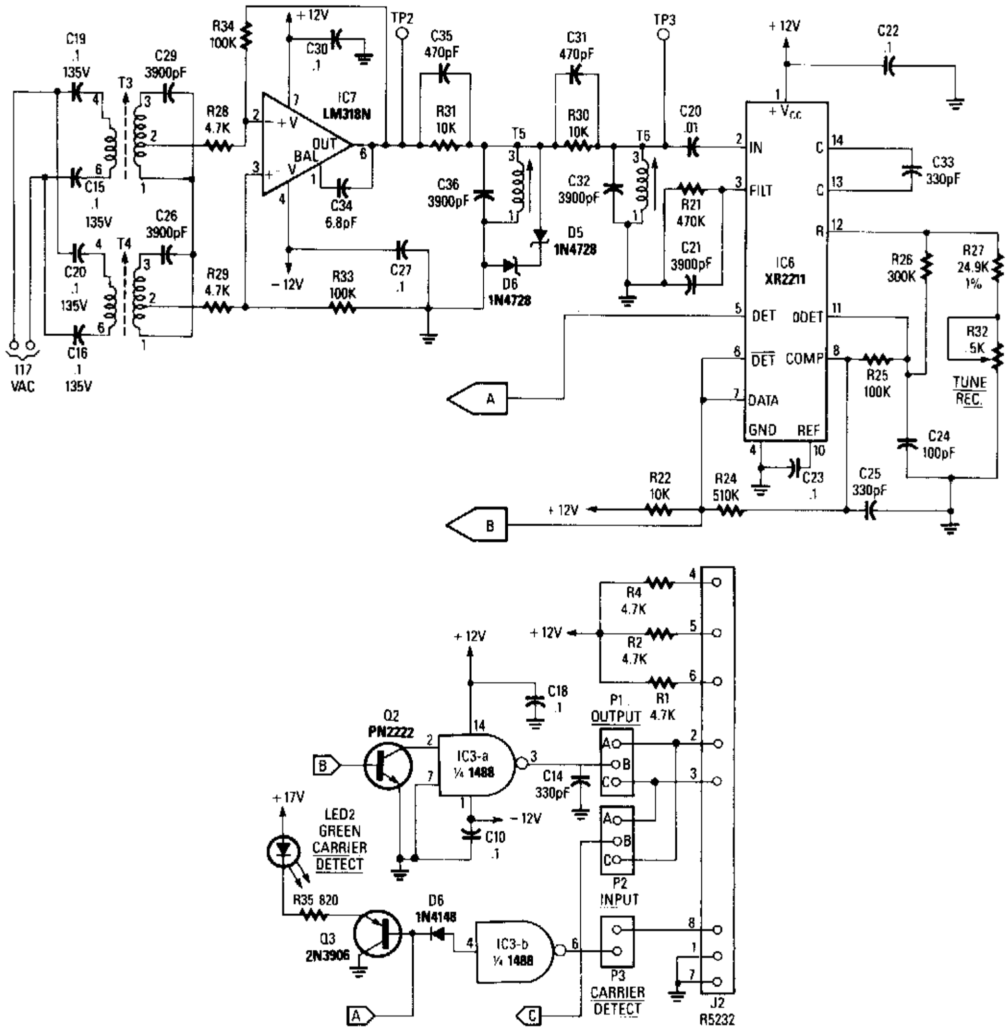
$$f_2 = f_1 + \Delta f_1$$

where:

$$\Delta f_1 = \frac{1}{R_4 C_1}$$

In this circuit, data at input C is amplified by IC2 and then fed to modulator IC1. IC1 generates two frequencies, depending on the values of C_5, R_{10}, R_3, R_{11}, and R_6. The frequency f_1 is generated if pin 9 IC1 is low and f_2 if pin 9 IC1 is high. A square wave appears at pin 13 of IC1 and is fed to Q1, an amplifier stage, that is coupled via tuned transformer T, to the ac line via C1 and C2. Notice that, for safety reasons, C1 and C2 must be specifically rated for the ac line voltage.

CARRIER-CURRENT RECEIVER FOR DATA TRANSMISSION



RADIO-ELECTRONICS *Fig. 18-2*

$$\text{Receive frequency: } f_0 = \frac{1}{(R_{27}+R_{32})\ C_{33}}$$

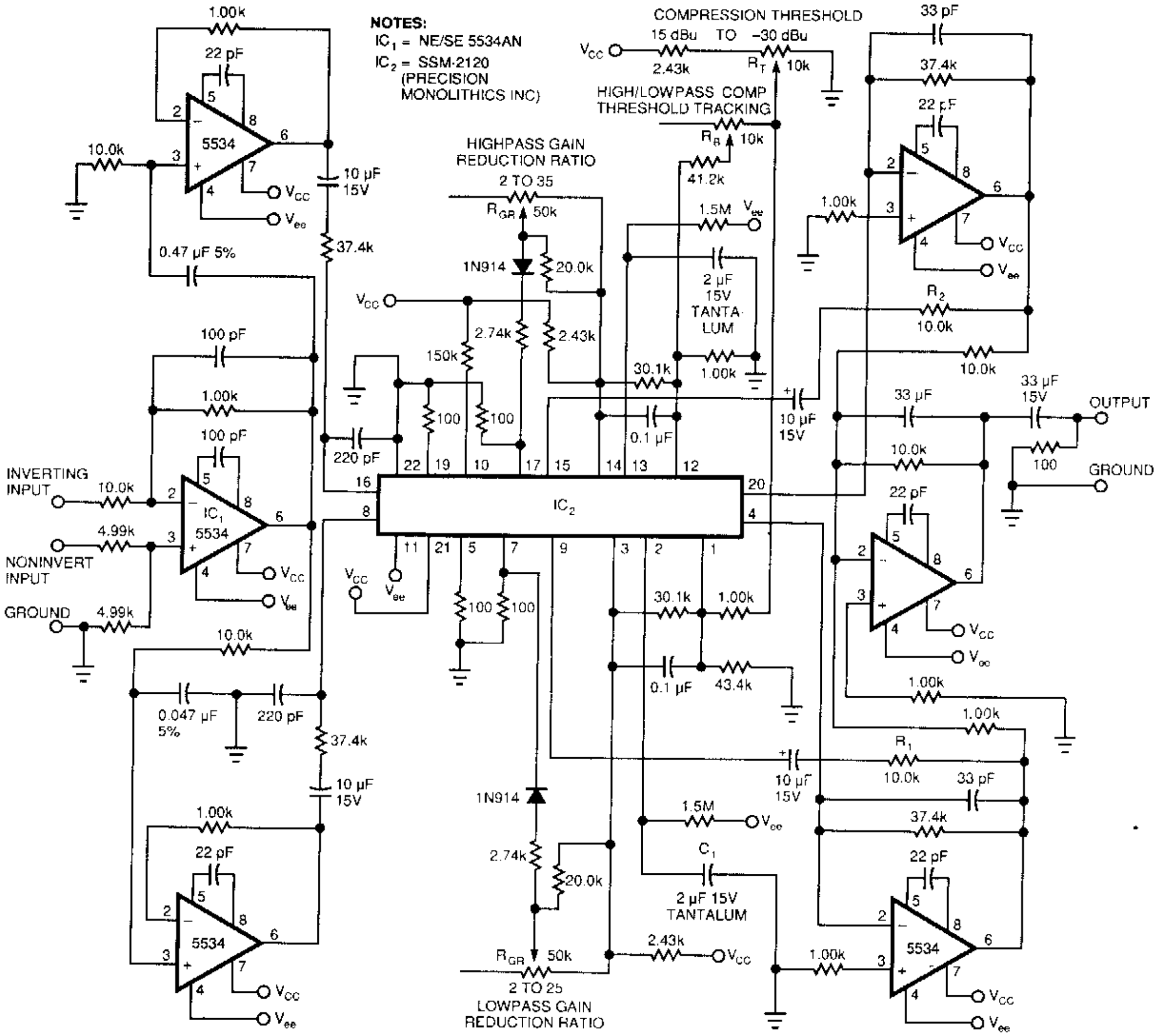
This receiver consists of an input network amplifier IC7 FSK PLL detector ICG, and output amplifier/interface Q2, Q3, IC3A and IC3B, a 1488 Quad RS232 line driver of the carrier-current signal. Tuned amplifier IC7 amplifies this signal and drives PLL detector IC8. The values shown in the circuit are suitable for operation in the 100-kHz range. Recovered data at pins 5, 6, 7 is fed to the output amplifier/interface circuit (Fig. 6). This circuit is also used with the carrier-current data transmitter to form a pair.

19

Compressor/Expander Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 663. The figure number in the box of each circuit correlates to the entry in the Sources section.

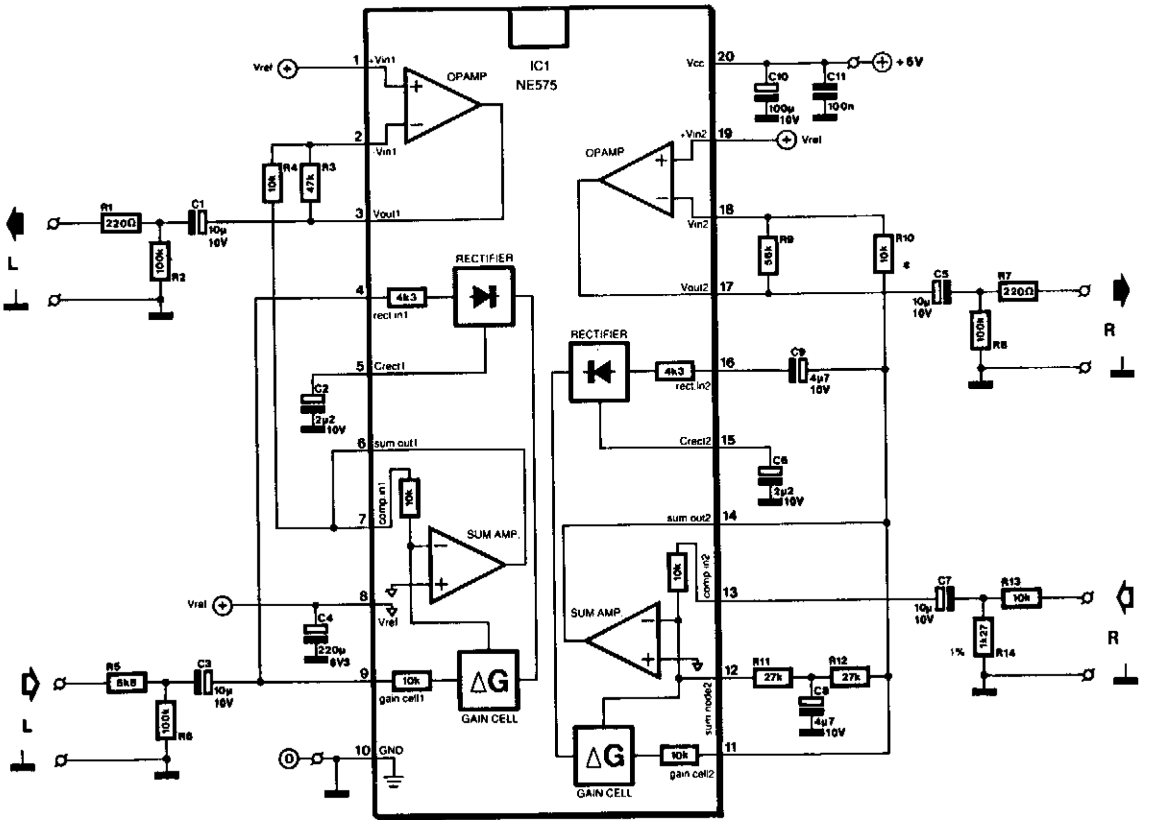
AUDIO COMPRESSOR/AUDIO-BAND SPLITTER



EDN *Fig. 19-1*

This 2-band compressor splits the audio into high and low frequencies and allows independent adjustment of each. Two active filters drive the two halves of dual voltage controlled amplifier/rectifier IC. Each section has a dynamic range greater than 100 dB. Compression gain slopes are adjustable from 2 to 25 for both audio bands. R_B adjusts the threshold amplitude between the two bands. RK1 and R2/C2 control the compressor attack times (10 kΩ and 2 μF, respectively), while the 1.5-MΩ resistor in the integrator circuit controls the release line.

UNIVERSAL COMPANDER



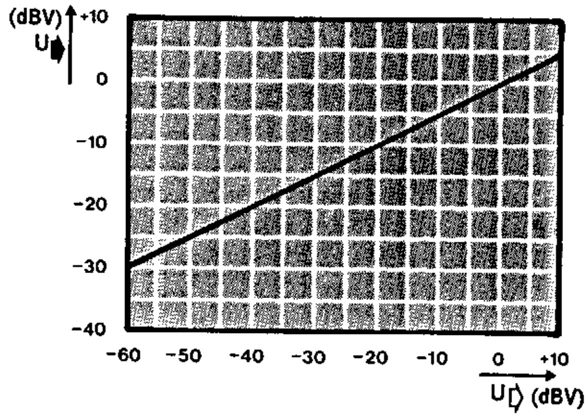
ELEKTOR ELECTRONICS

Fig. 19-2(a)

Signetics' type NE575 compander IC is intended primarily for use with battery power supplies of 3 to 7 V (max. 8 V). It draws a current of 3.5 mA at 3 V and 5 mA at 7 V. The compander process (compression at the input, expansion at the output) significantly improves the signal-to-noise ratio in a communications link.

The IC contains two almost identical circuits, of which one (pins 1 to 9) is arranged as an expander. The other (pins 11 to 19) can be used as expander, compressor or automatic load control (ALC), depending on the externally connected circuit. For the compressor function, the inverting output of the internal summing amplifier is brought out to pin 12. This is not the case in the expander section, where a reference voltage is available at pin 8. This pin is interlinked to pins 1 and 19 to enable the setting of the dc operating point of the op amps.

The op amp in the expander section, pins 1 through 3, serves as output buffer in the compressor section, pins 17 through 19 as the input buffer. The IC has a relatively high output sensitivity and is evidently intended for processing small signals (microphone output level). A signal of 100 mV, for instance, is amplified by 1 only. The present circuit caters to larger input signals (line level); its maximum input level is 1.5 Vrms.



ELEKTOR ELECTRONICS *Fig. 19-2(b)*

With a 1-V input into R13, a potential of about 500 mV exists between compressor output R7 and expander input R5. The compression characteristic is shown in Fig. 19-2(b). The signal range is reduced by about one half at the output, which is doubled in the expander. Thus, the range after compression and expansion is the same again, but that is not necessarily the case with the input and output level. The compander can be arranged to provide a constant attenuation or amplification. With the circuit values as shown in the diagram, the input and output levels are the same. The prototype had an overall gain of 0.5 dB when the expander input was connected directly to the compressor output.

To allow acceptance of high input levels, R13, R14, and the compressor input resistance form a 10:1 attenuator. At the expander input, R5 and the expander input impedance of about 3 kΩ form a potential divider. If the compander is to be used with smaller signals, the attenuation can be reduced as appropriate. If the input level lies below 100 mV, R5, R13 and R14 can be omitted.

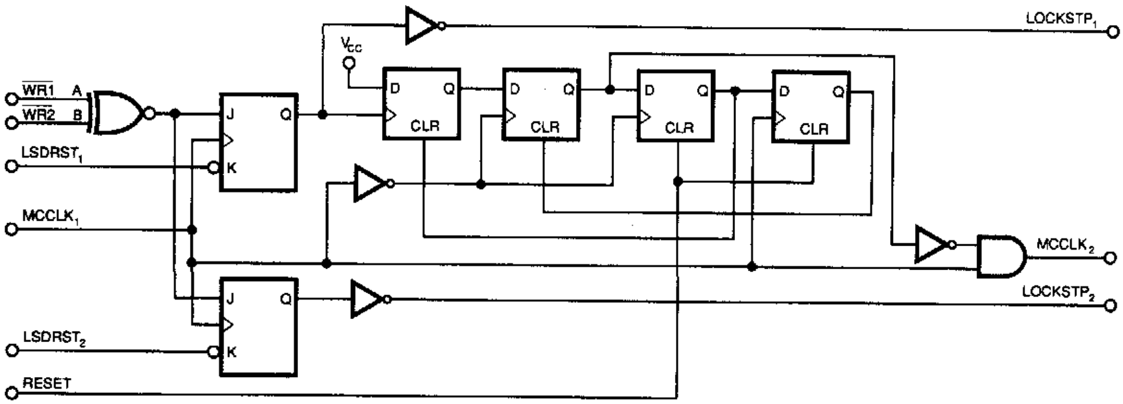
The compander covers the frequency range of 20 Hz to 20 kHz, the overall distortion is less than 1%, and the signal-to-noise ratio is about 80 dB.

20

Computer-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 663. The figure number in the box of each circuit correlates to the entry in the Sources section.

DUAL 8051s EXECUTE IN LOCK-STEP



EDN

Fig. 20-1

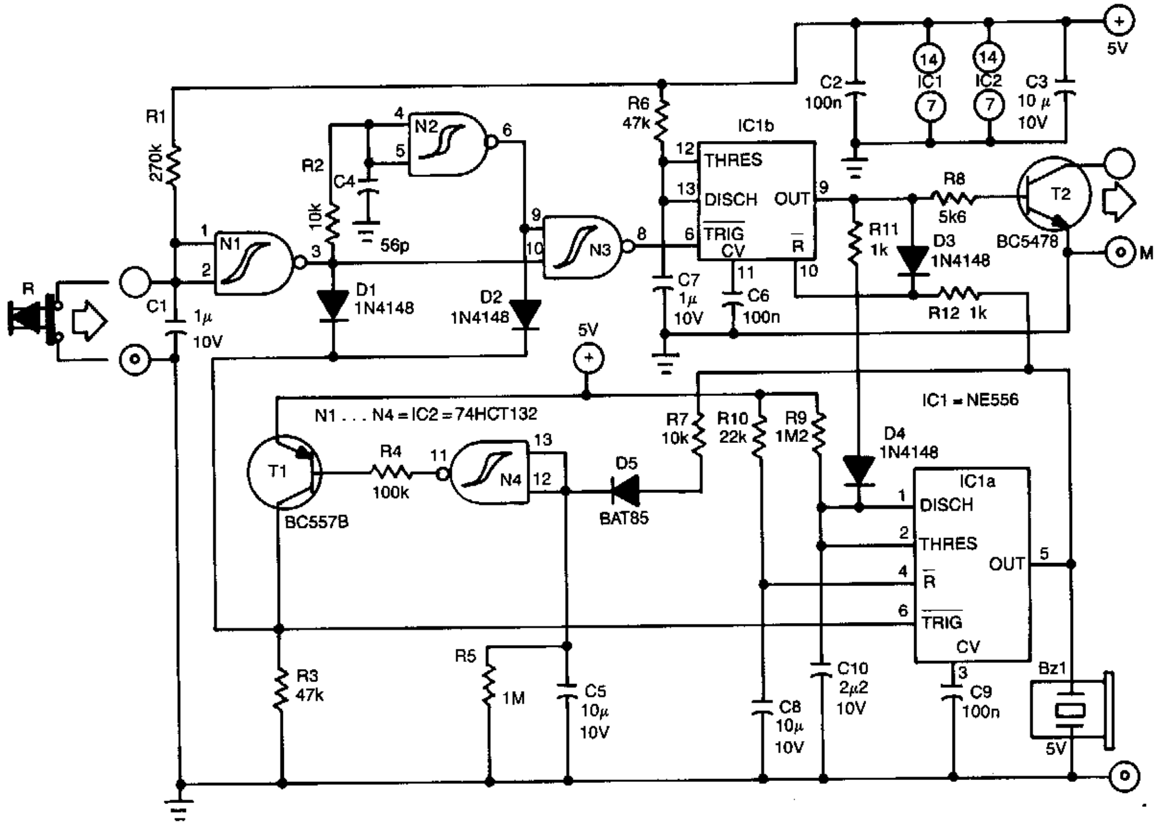
Table 20-1. Synchronizing firmware loop

```
SYNCUP:   MOV    P3, #OFFH        ;Activate internal port
;                                    pull-ups.
          MOV    DPTR, #SAFLOC    ;Location for safely writing
;                                    values so that WR*
                                     can be generated.
          MOV    A, #SAFVAL       ;Value to be written into
;                                    location DPTR is pointing
;                                    to. Use value that will
;                                    not cause adverse effects.
WAIT:     CLR    P1.0             ;Toggle the JK flip flop
          SETB   P1.0               lock-step-reset (K input).
          MOVX   @DPTR, A         ;Generate WR*.
          JNB    P3.5, WAIT       ;If signal is high (LOCKSTP1)
;                                   or LOCKSTP2) then lock
;                                   step has been achieved.
;                                   Otherwise, try again.
;         Program will now execute synchronously on each
;         microcontroller.
```

This hardware-software combination deletes clocks from the slave μP until both μPs synchronize. The firmware loop causes each μP to generate a WR signal once per loop. The circuit exclusive-ORs the two WR signals to produce a miss-compare pulse. The miss-compare pulse latches into the two JK flip-flops via outputs LOCKSTP1 and LOCKSTP2. A high on these signals indicates that the μPs are in lock-step, causing both μPs' programs' execution to exit the firmware loop. If you use discrete components, you'll probably want to use the *Q* output of the JK flip-flop and delete the circuit's inverters.

The listing uses the μPs' ports 1 and 3. You cannot use a memory-mapped location for the lock-step-detect clear (K input) because this scheme would generate additional WR signals. You could apply this idea to other μPs, perhaps using their RD signals. This way, generating an RD signal to activate the lock-step-detect clear would not affect the synchronization inputs.

RESET PROTECTION FOR COMPUTERS



ELEKTOR ELECTRONICS

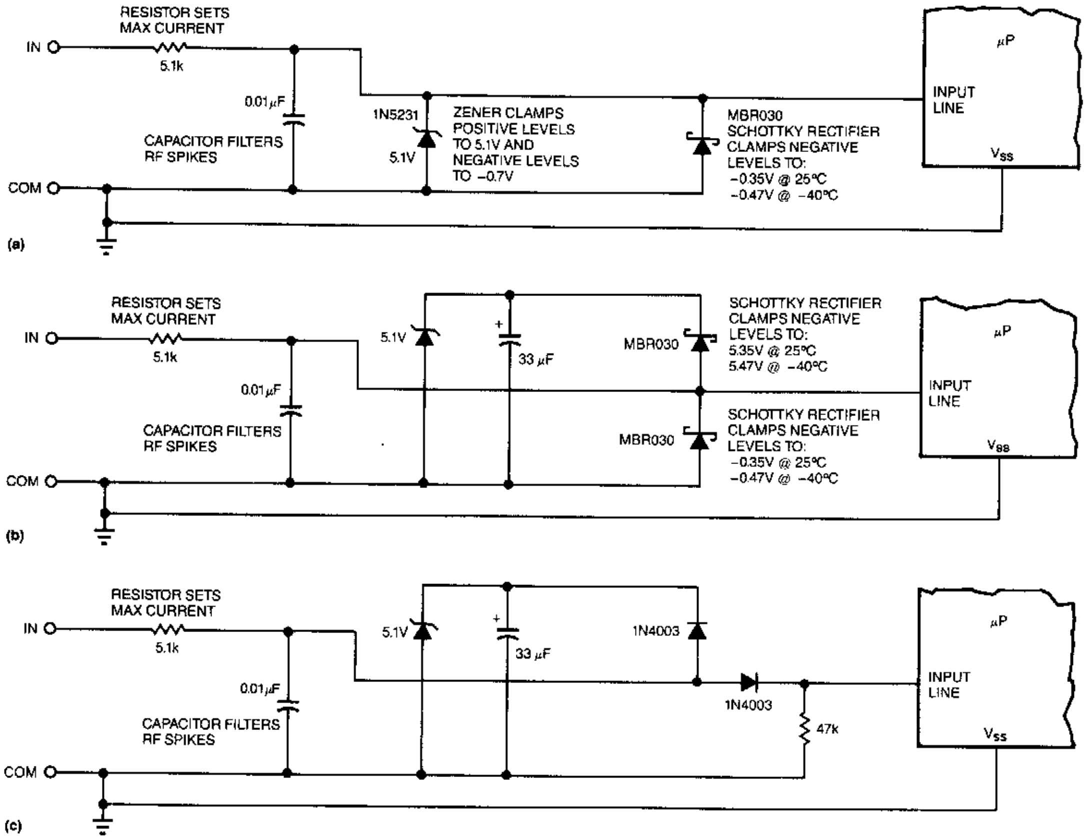
Fig. 20-2

This protection circuit is inserted between the reset switch and the motherboard. The earth connection of the computer must be linked to terminal M of the protection circuit. The protection circuit can draw its power from the computer supply.

When the circuit has been fitted, operation of the reset switch will not immediately restart the computer. Instead, a buzzer will sound to alert you to the reset operation. The buzzer is actuated for 4 s by monostable IC1A, which is triggered by the reset switch. During these 4 s, the output, pin 5, of IC1A ensures that the reset function, pin 10, of IC1B is disabled. When the reset switch is operated again, monostable IC1B will be triggered and this starts the reset procedure. Transistor T2 is then switched on for 0.5 s and the buzzer is deactuated via R11 and D4.

The circuit around T1 and N4 ensures that IC1A can accept trigger pulses again 10 s after the mono time of IC1B has lapsed. This arrangement prevents, for example, children operating the reset switch.

3 μP I/O LINE PROTECTORS



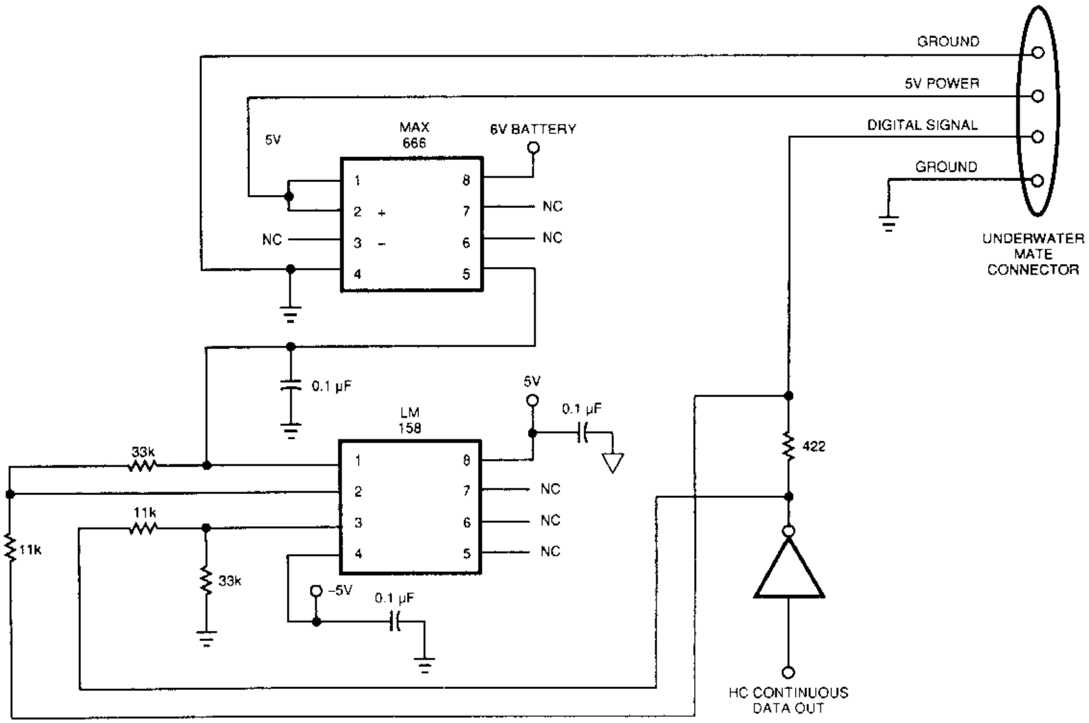
EDN

Fig. 20-3

In Fig. 20-3(a), a 5.1-V zener diode clamps positive-going transients, and a Schottky rectifier clamps negative-going transients. The Schottky rectifier has problems at both ends of the temperature scale. At 125°C (257°F), its leakage current can reach 50 μA when the input line is at 5 V. This leakage is not a big deal unless the input resistor has a value of 100 kΩ or more. More troubling, at temperatures below −40°C (−40°F), the Schottky rectifier's forward voltage rises to about 0.47 V, which is perilously close to the −0.50-V max spec that most HCMOS-type μP's inputs can tolerate.

The third circuit, Fig. 20-3(c), uses two regular silicon rectifiers. One rectifier is connected in series with the input line, thereby isolating the μP's inputs from negative-going voltage spikes. The other rectifier is in series with a 5.1-V zener, which clamps positive-going transients.

DATA LINE REMOTE SHORT SENSOR



EDN

Fig. 20-4

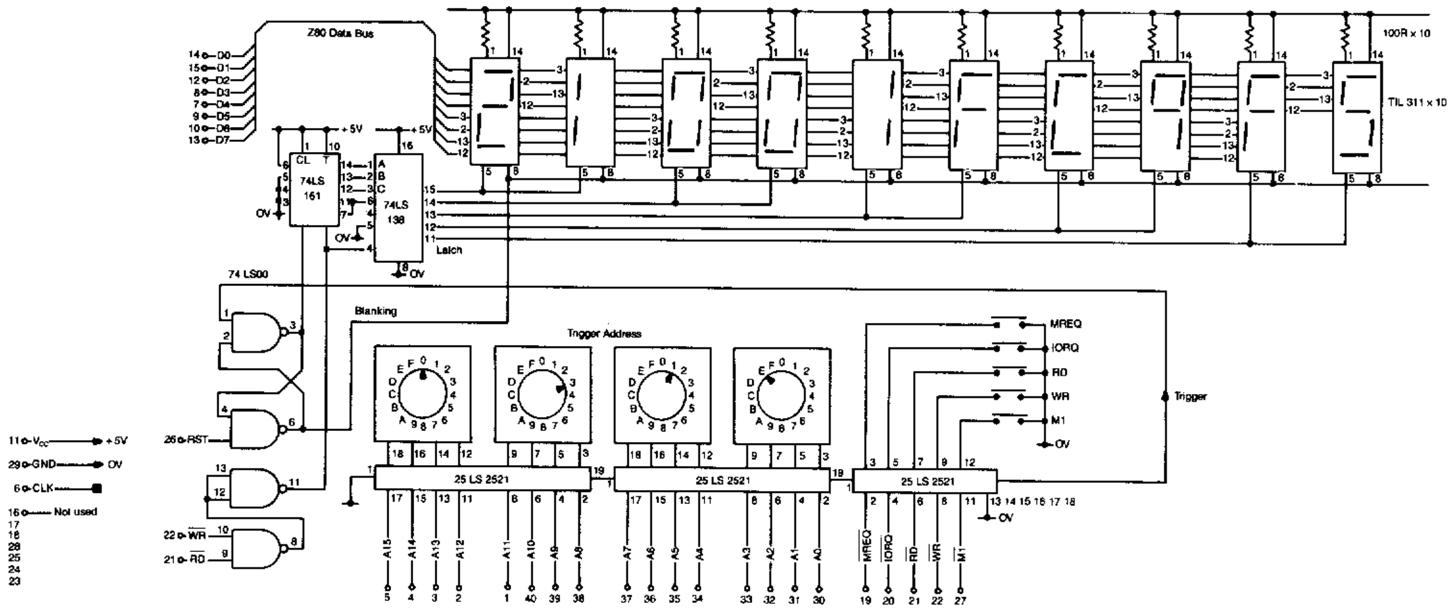
Sensing short circuits in equipment that performs under water is especially critical, but Fig. 20-4's wet-mate connector design also suits other remote short-circuit sensing applications. Because of the limits imposed by the battery and voltage levels, the circuit uses the data line to sense short circuits.

The differential voltage that develops across the 422-Ω resistor in the data line drives a low-bandwidth op amp, which amplifies and filters the differential signal. The resistor values produce a gain of 3. The op amp's output controls the voltage regulator's shutdown pin.

To operate correctly, the circuit must have a continuous stream of digital data. Under normal conditions, and using high-speed CMOS logic, the data source sinks less than 10 µA. This normal operation generates about −3 mV across the sense resistor. The op amp's output will be slightly negative, producing a solid ON signal to the voltage-regulator chip. When a short occurs, the resistor and op amp together produce an average of 2.4 Vdc. This voltage provides a solid OFF to the voltage-regulator chip.

The peak signal-line current is about 12 mA (5-V data divided by 422 Ω), which HCMOS logic can handle. The addition of the resistor and op amp only changes the rise time to about 40 ns and doesn't cause any problems with the 2.5-MHz data rate. When the short is no longer present, the voltage regulator chip turns on again. You can use the same circuit with any TTL on/off-type voltage-regulator IC.

Z-80 BUS MONITOR/DEBUGGER



ELECTRONIC ENGINEERING

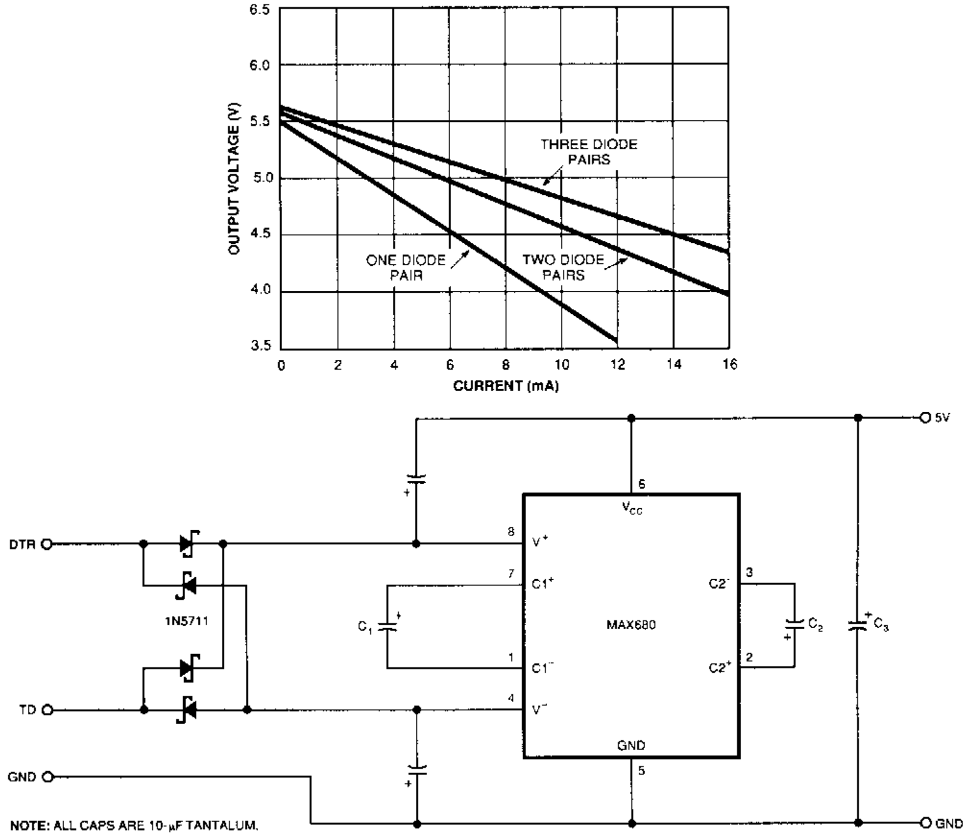
Fig. 20-5

Getting microprocessor designs to work is notoriously difficult when both the software and hardware are new. The usual approach is to run test routines that address memory and I/O, but do not rely on their correct functioning. However, miswiring in any part of the circuit usually leads to a misleading jumble of signals that might require a logic analyzer to interpret.

This simple circuit will trace the program execution and help point to the problems. Although the circuit shows connections for a Z-80, the circuit can very easily be adapted for any 8-bit microprocessor or with additional circuitry for CPUs of any bus width.

The circuit consists of a 5-byte hexadecimal display and comparator, which are wired to a 40-pin IC test clip. The test clip sits over the microprocessor (in this case a Z-80), where it gets power and all the required signals. The address bus and control lines are fed to the comparator, where (by means of switches) a trigger condition can be set. Following the trigger, the next 5 occurrences of either RD or WR will latch the contents of the data bus into the 5 hex displays, each in turn.

For example, select address 0000 M1 Rd and reset the CPU. The displays will show the very first instruction fetch, followed by its data and any consequent action. Even details, such as stack writes and subroutine addresses, are included. To trace longer portions of a program the address switches can be incremented to follow the execution path.

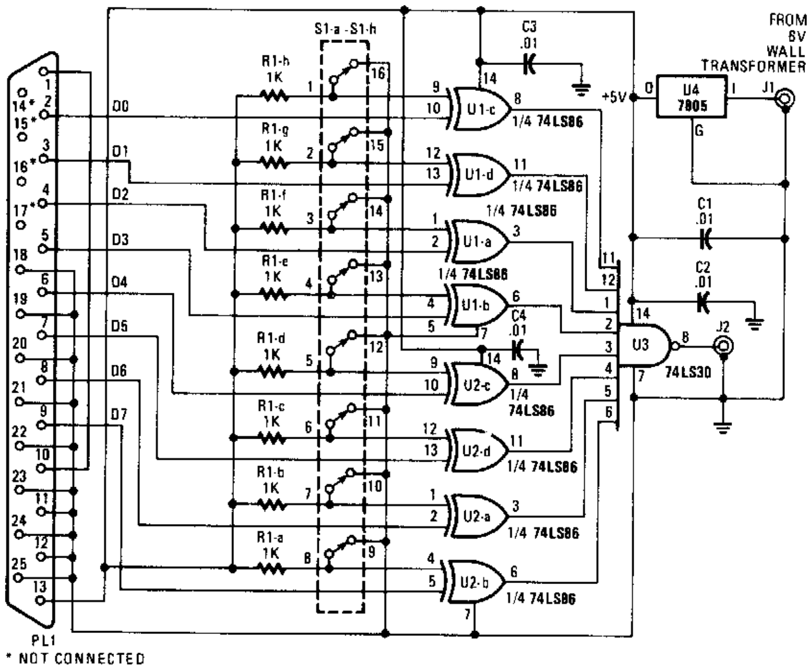


EDN **Fig. 20-6**

The circuit illustrates a way to power CMOS ICs from RS-232C lines. The MAX680 is normally used to generate a voltage equal to $\pm 2 \times V_{CC}$. This circuit does exactly the opposite. It takes in ± 10.5 to ± 12 V on the DTR and TD lines and puts out a 5.25- to 6-V signal. A pair of Schottky or silicon diodes rectifies each RS-232C line. The resultant energy is stored by the capacitors attached to the IC's V+ and V− pins. C1 or both C1 and C2 then reverse-charge pump the energy stored at the V+ and V− pins to C3. The input source current of the MAX680 is approximately equal to the voltage drop of any one of the diodes that is divided by the series resistance of 160 Ω. When you drive this circuit from a 1488 driver with a ± 12-V supply, it can deliver 5 V at 3 mA.

To increase the output current, you can use as many as three sets of diodes on each RS-242 line to provide 5 V at 8 mA. The more diodes you use, the lower the source resistance: R_S equals the inversion of the sum of the diodes' conductances. If your circuit requires even more output current, you can place two MAX680s in parallel, tie their V+ and V− capacitors together, and use separate C1 and C2 capacitors for each ship. If you do connect the devices in parallel, make sure not to exceed the power capability of the RS-232C lines.

BIT GRABBER

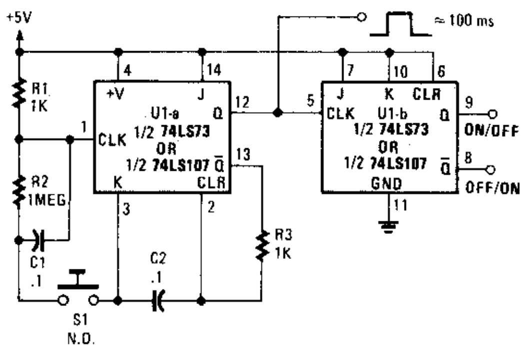


POPULAR ELECTRONICS

Fig. 20-7

When the user set character (D_0 through D_7) from a computer matches the character programmed on S1A through S1H, the output from J2 becomes low. This device can be used as a test aid to check printer cable or as a control circuit for interfacing with a computer.

SWITCH DEBOUNCER

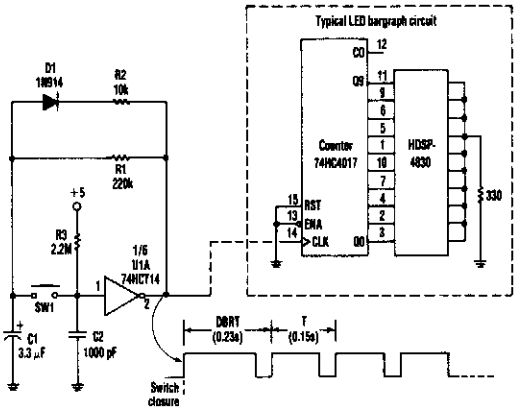


Using a 7473 JK flip-flop U1A connected as a monostable to drive U1B, as a switch debouncer. The circuit is self-clearing during power up. A 100-ms pulse is available at pin 12 U1A.

POPULAR ELECTRONICS

Fig. 20-8

SWITCH DEBOUNCER WITH AUTO REPEAT



ELECTRONIC DESIGN *Fig. 20-9*

This circuit produces an output pulse when SW1 (pushbutton) is depressed. It also becomes a hysteresis gate oscillator. D1 and R2 add asymmetry. The DBRT (*delay before repeat time*) is caused by the oscillator start-up conditions: C1 has to change from zero to the upper gate threshold rather than to the lower threshold.

The auto repeat time:

$$t = \frac{(R_1 + R_2)(C_1 + C_2)(\text{Gate Hysteresis})}{V_{\text{SUPPLY}}}$$

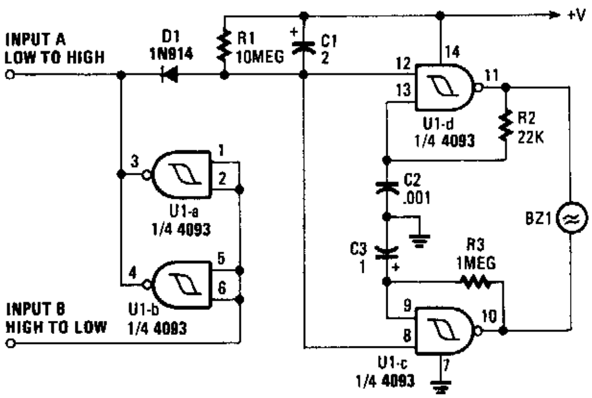
Gate hysteresis ≈ 1 V for 74HCT14 gate

DBRT = 0.7T (upper gate threshold hysteresis)

Upper gate threshold ≈ 2.3 V for HCT14

$$R_1 << R_3$$
$$R_2 << R_1$$

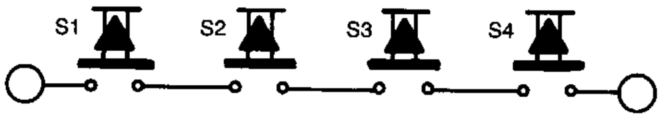
PRINTER ERROR ALARM



POPULAR ELECTRONICS *Fig. 20-10*

When a printer is shut down, this alarm sounds an alarm. The input can be either a high-to-low or low-to-high transition. This can be a logic level that corresponds with the printer being on or off. The oscillator produces an interrupted (on-off) tone.

CHILD-PROOF RESET SWITCH



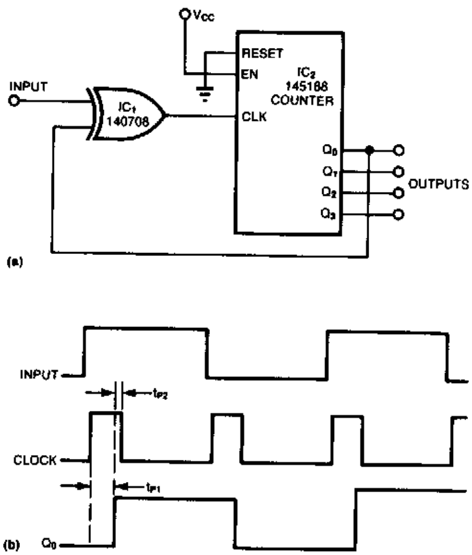
ELEKTOR ELECTRONICS *Fig. 20-11*

The reset switch on a computer is very important. If an operating instruction threatens to wreck the internal management of a computer, the reset button is often the only way of avoiding a possible disaster. On the other hand, it also could cause a disaster.

It is particularly important that children or pets cannot inadvertently operate the control. The circuit proposed here should put an end to your worries in this respect. Instead of one reset switch, it is necessary to press four switches simultaneously. The chances of this happening via accident, child, or pet are negligible.

The four switches are placed in positions that make it impossible to operate them all with one hand. Instead, two of them can be operated with the fingers of one hand and the other two with the fingers on the other hand. As shown, the four switches are connected in series and are intended to replace the existing switch.

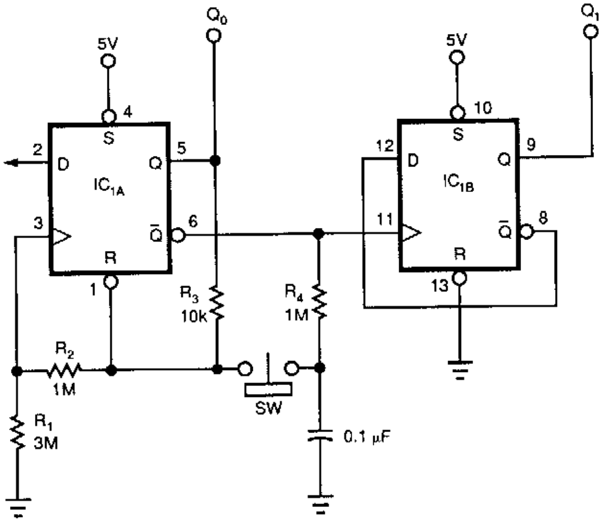
XOR GATE



Inverting the negative-going input transactions allows the counter to count both positive- and negative-going edges. The XOR gate transforms the input signal into a series of short pulses whose width is equal to the sum of the counter and gate propagation delays.

EDN *Fig. 20-12*

FLIP-FLOP DEBOUNCER SWITCH



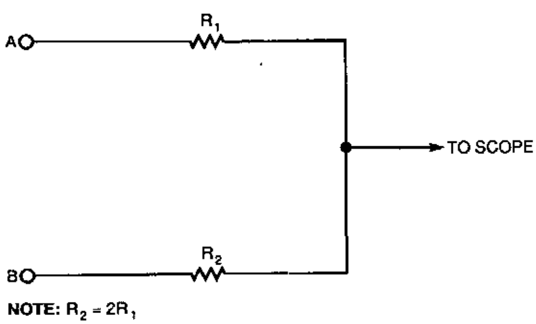
EDN

Fig. 20-13

Although this circuit uses a 74HC74, any CMOS variant of this flip-flop will work. IC1A acts as a true/complement buffer. R1 and R2 ensure that IC1A comes out of reset before the clock's edge occurs. R3 applies IC1A's logic state to pins 1 and 3. When the switch closes, the next logic state stored on the capacitor transfers to the flip-flop's reset and clock inputs. Releasing the switch lets the capacitor charge to the next state via R4. IC1A's output is the LSB; IC1B's output is the MSB.

Notice that the counter's state advances when the switch is first pressed, rather than when it's released; the latter is the case with many other switch-debouncing schemes. You can replace R1 with a 22-pF capacitor to reduce the circuit's sensitivity to parasitic effects. The addition of this capacitor also lets you lower the magnitude of R2 and R3 by a factor of 10.

DIGITAL LEVELS SCOPE DISPLAYS 2 LOGIC SIGNALS ON 1 SCOPE



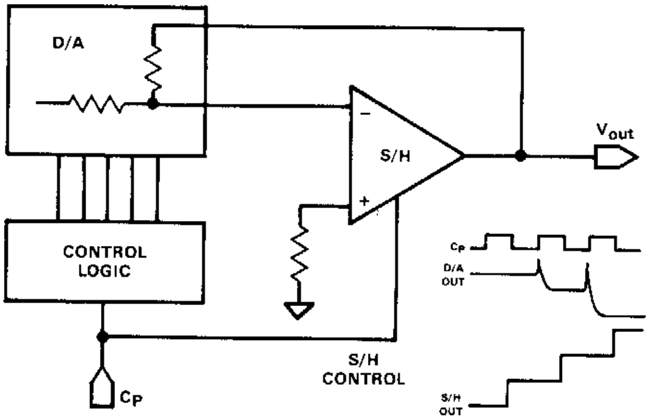
EDN

Fig. 20-14

Using this simple resistor circuit, you can trick your oscilloscope into displaying two logic signals on one channel. If you select R_2 to be twice R_1, the scope trace will show one of four distinct analog levels for each possible combination of the states of inputs A and B.

Of course, the voltage levels that your oscilloscope sees depends heavily on the current-sourcing capability of your digital logic. Because you must use high resistances, this technique has a limited frequency range.

DEGLITCHER



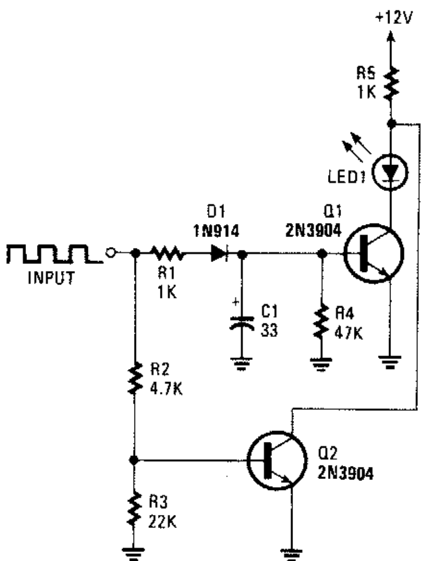
HARRIS

Fig. 20-15

Glitch has been a universal slang expression among electronics people for an unwanted transient condition. In D/A converters, the word has achieved semiofficial status for an output transient, which occurs when the digital input addressed is changed. The sample/hold amplifier does double duty, serving as a buffer amplifier as well as a glitch remover, delaying the output by 1/2-clock cycle.

The sample/hold can be used to remove many other types of glitches in a system. If a delayed sample pulse is required, it can be generated using a dual monostable multivibrator IC.

STALLED-OUTPUT DETECTOR



This circuit can be used to detect a stuck output or node in a circuit, or a loss of data or pulses. The pulse train charges C1 and biases Q1 on, which lights the LED. If the input remains high, Q2 extinguishes the LED.

POPULAR ELECTRONICS

Fig. 20-16

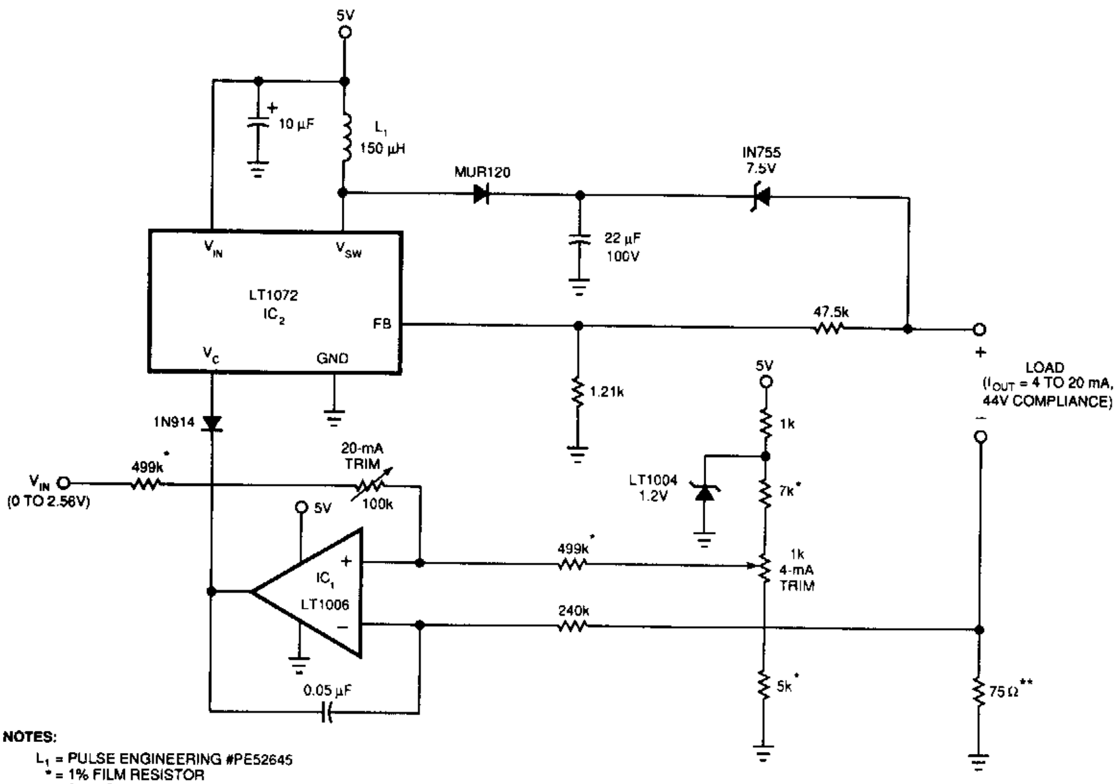
21

Converters

The sources of the following circuits are contained in the Sources section, which begins on page 664. The figure number in the box of each circuit correlates to the entry in the Sources section.

4-to-20-mA Current Loop
Power-Saving Intermittent Converter
Current-to-Frequency Counter
Sawtooth Converter
Period-to-Voltage Converter
Rectangular/Triangular Waveform Converter
μP-Controlled Negative Voltage Converter
dc/dc Converter
Voltage-to-Current Converter
1-to-5-V dc/dc Converter
9-to-5-V Converter
+3-V Battery to +5-V dc/dc Converter
Sine-Wave/Square-Wave Converter
Precision Full-Wave ac/dc Converter

4-TO-20-mA CURRENT LOOP



NOTES:

L1 = PULSE ENGINEERING #PE52645
* = 1% FILM RESISTOR
** = 0.1% FILM RESISTOR
RECOMMEND CERMET POTENTIOMETERS

EDN

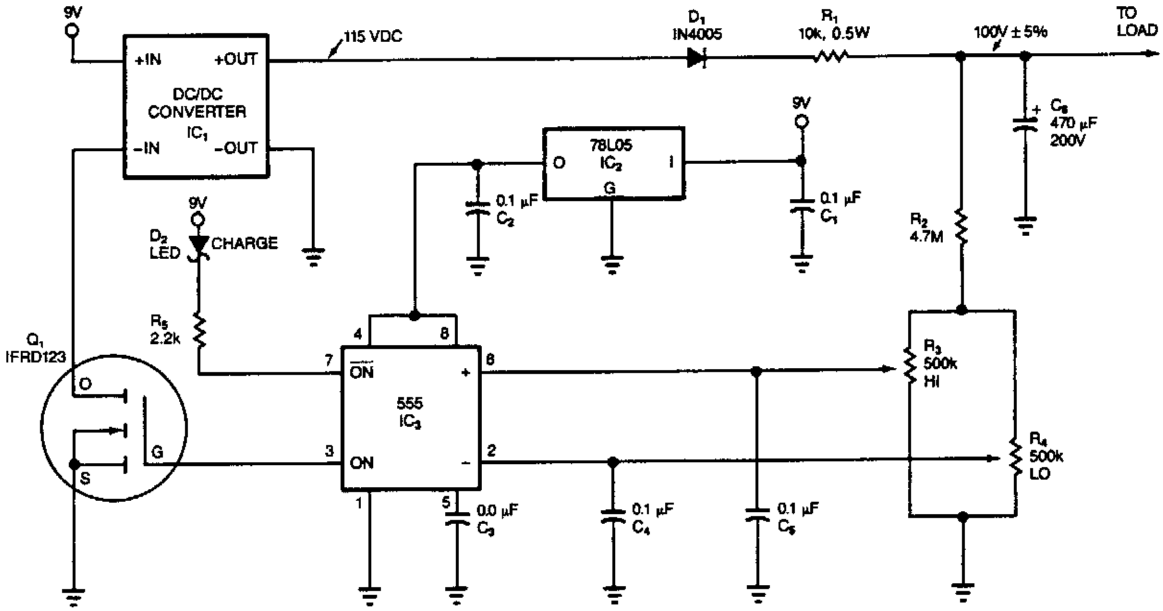
Fig. 21-1

This 5-V circuit utilizes a servo-controlled dc/dc converter to generate the compliance voltage necessary for loop-current requirements. This circuit will drive 4 to 20 mA into loads as high as 2 200 Ω with 44 V of compliance. It is inherently short-circuit protected. A current source by definition limits current regardless of the load.

The circuit's input voltage and the 4-mA trim network determine IC1's positive input voltage. IC1's output biases the LT1072 switching regulator's V_C pin. The resistors connected to the regulator's feedback pin, FB, prevent the circuit output from running away in the event that the load opens up.

Normally, IC1 controls the loop. However, if the load opens, IC1 receives no feedback. Under this condition, the FB pin becomes active when it equals 1.2 V and forces the loop to close locally around the regulator by activating IC2's internal amplifier. Thus, the circuit automatically changes from a current to a voltage regulator, thereby preventing excessive output voltages.

POWER-SAVING INTERMITTENT CONVERTER



EDN

Fig. 21-2

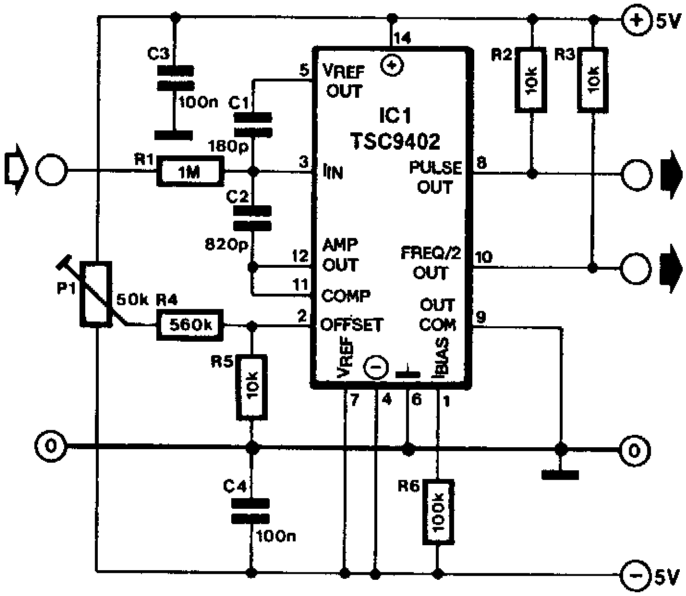
This circuit switches its dc/dc converter, IC1, off whenever the large filter capacitor, C6, has sufficient charge to power the load. This particular circuit uses a dc/dc converter that produces 115 Vdc from a 9-Vdc input; you can tailor the circuit to suit other converters. The heart of the circuit is a 555 timer configured as a dual-limit comparator. Thus, the 555 turns the converter on or off, depending on the voltage across C6. The 555's complementary output lights the charge LED when the FET is on.

Initially, the voltage on C6 is zero, and the 555's output turns on the FET, Q1, in turn, enabling the converter to run, which charges C6. When the voltage on the capacitor reaches the value set by R3, the 555 turns the converter off. Then, C6 slowly discharges into the combined load of the voltage divider (R2, R3, and R4) and the reverse-biased blocking diode, D1.

When the voltage falls below $1/3\ V_{CC}$, the 555 restarts the dc/dc converter. If this circuit powers a load that periodically goes into a zero-power, shutdown mode, the 555 switches the dc/dc converter on full time whenever the load kicks in.

When the supply voltage falls below 7.5 V, the output of the converter is no longer high enough to charge, the LED doesn't light. The circuit uses 205 mA when the converter is on and 10 mA when the converter is off. The duty cycle comprises a 5-s ON period, a 150-s OFF period, and it represents a 92% power reduction. You can further reduce power consumption by removing the charge LED and using a CMOS 555 and a CMOS 78L05 regulator.

CURRENT-TO-FREQUENCY CONVERTER



ELEKTOR ELECTRONICS

Fig. 21-3

Teledyne Semiconductor's Type TSC9402 IC is eminently suitable as an inexpensive current-to-frequency converter. The maximum input current of the design shown in the diagram is 10 μA (input voltage range is 10 mV to 10 V), while the output frequency range extends from 10 Hz to 10 kHz. The conversion factor is exactly 1 kHz/μA. The factor can be altered by changing the value of R1—as long as the maximum input current of 10 μA is not exceeded.

The circuit has two outputs. That at pin 8 is a short-duration pulse, whose rate is directly proportional to the input current; that at pin 10 is a square wave of half the frequency of the pulse at pin 8.

Calibrating the circuit is fairly simple. Connect a frequency meter to pin 8 (preferably one that can read tenths of a hertz) and connect a voltage of exactly 10 mV to the input (check with an accurate millivoltmeter). Adjust P1 to obtain an output of exactly 10 Hz. Next, connect a signal of exactly 10 V to the input and check that the output signal has a frequency of 10 kHz. If this frequency cannot be attained, shunt C1 with a small trimmer or replace R1 by a resistor of 820 kΩ and a preset of 250 kΩ.

The circuit may be adapted to individual requirements with the aid of:

$$f_{out} = I_{in} U_r (C_1 + 12 \text{ pF}) \qquad [\text{Hz}]$$

The reference voltage, U_r, here is -5 V.

SAWTOOTH CONVERTER

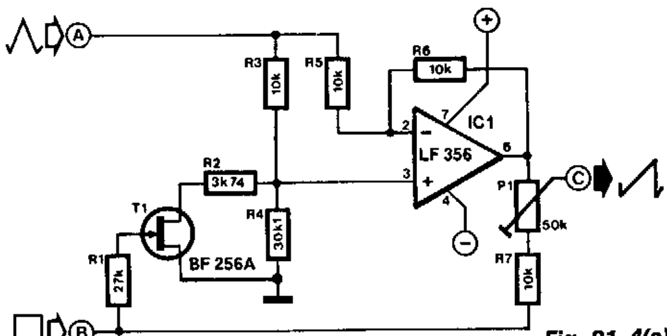


Fig. 21-4(a)

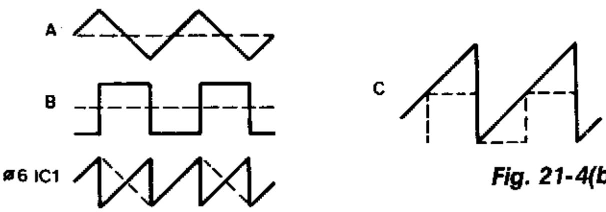


Fig. 21-4(b)

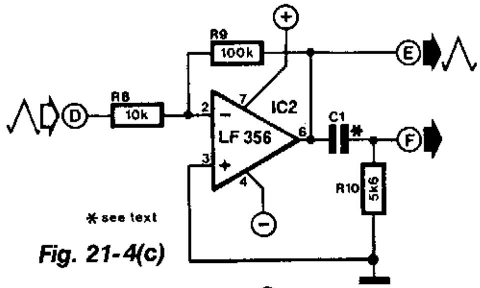


Fig. 21-4(c)

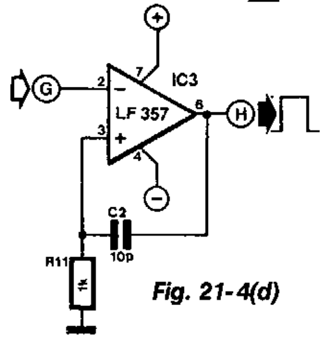


Fig. 21-4(d)

ELEKTOR ELECTRONICS

Simple function generators normally provide sinusoidal, rectangular, and triangular waveforms, but seldom a sawtooth. The circuit in Fig. 21-4(a) derives a sawtooth signal from a rectangular and triangular signal. Its quality depends on the linearity of the triangular signal, the slope of the edges of the rectangular signal and the phase relationship between the rectangular and triangular signals.

The conversion is carried out in IC1. Whether the triangular signal at input A is converted or not by IC1 depends on the state of T1. This FET is controlled by the rectangular signal at input B.

The signal at the output of the op amp is a sawtooth (see Fig. 21-4(b)) whose trailing edge is inverted. The frequency of this signal is double that of the input signals.

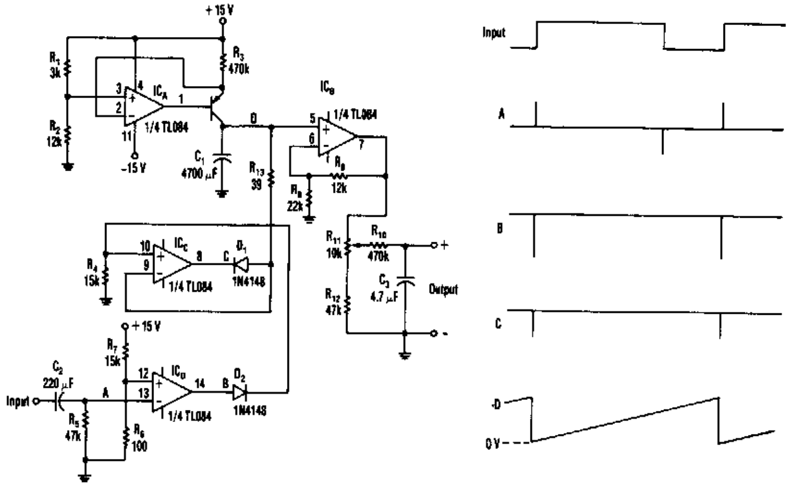
If in this state, the dc level of each inverted edge is raised sufficiently to make the lower level of that edge coincide with the higher level of the preceding edge, a sawtooth signal of the same frequency (but double the peak value of the input signals) is obtained. The dc level is raised by adding input B to the output of IC1 via R7 and P1. The preset should preferably be a multiturn type. Resistors R2 and R4 are 1% types.

If a rectangular signal is not available, or if its peak value is too small, the auxiliary circuits (shown in Figs. 21-4(c) and 21-4(d)) will be found useful. Figure 21-4(c) amplifies the triangular input at A by 10. Differentiating network C1/R10 derives rectangular pulses from the amplified triangular signal and these are available at F.

The pulses at F are shaped by the circuit in Fig. 21-4(d) to rectangular signals that have the same peak value as the supply voltage. Capacitor C2 increases the slope of the edge; it can be omitted for low-frequency signals.

The converter provides sawtooth signals over the frequency range of 15 Hz to 15 kHz. If the auxiliary circuits are used, capacitor C1 must be compatible with the frequency of the sawtooth signal (its value lies between 2 nF and 100 pF). The supply for all circuits can be between ±10 V and ±15 V. Each op amp draws a current of 4 to 6 mA.

PERIOD-TO-VOLTAGE CONVERTER



ELECTRONIC DESIGN

Fig. 21-5

ICA,R1, R2, R3, and Q1 form a current source. The current that charges C1 is given by:

$$I = \frac{V_{Dn} \times R_1}{(R_1 + R_2) \times R_3}$$

$$= \frac{15 \times 3\ \text{k}\Omega}{(3\ \text{k}\Omega + 12\ \text{k}\Omega) \times 470\ \text{k}\Omega}$$

$$= 6.4\ \mu\text{A}$$

The input signal drives ICD. Because ICD's positive input ($V+$) is slightly offset to $+0.1$ V, its steady state output will be around $+13$ V. This voltage is sent to ICC through D2, setting ICC's output to $+13$ V. Therefore, point D is cut off by D1, and C1 is charged by the current source. Assuming the initial voltage on C1 is zero, the maximum voltage (VC_{max}) is given by:

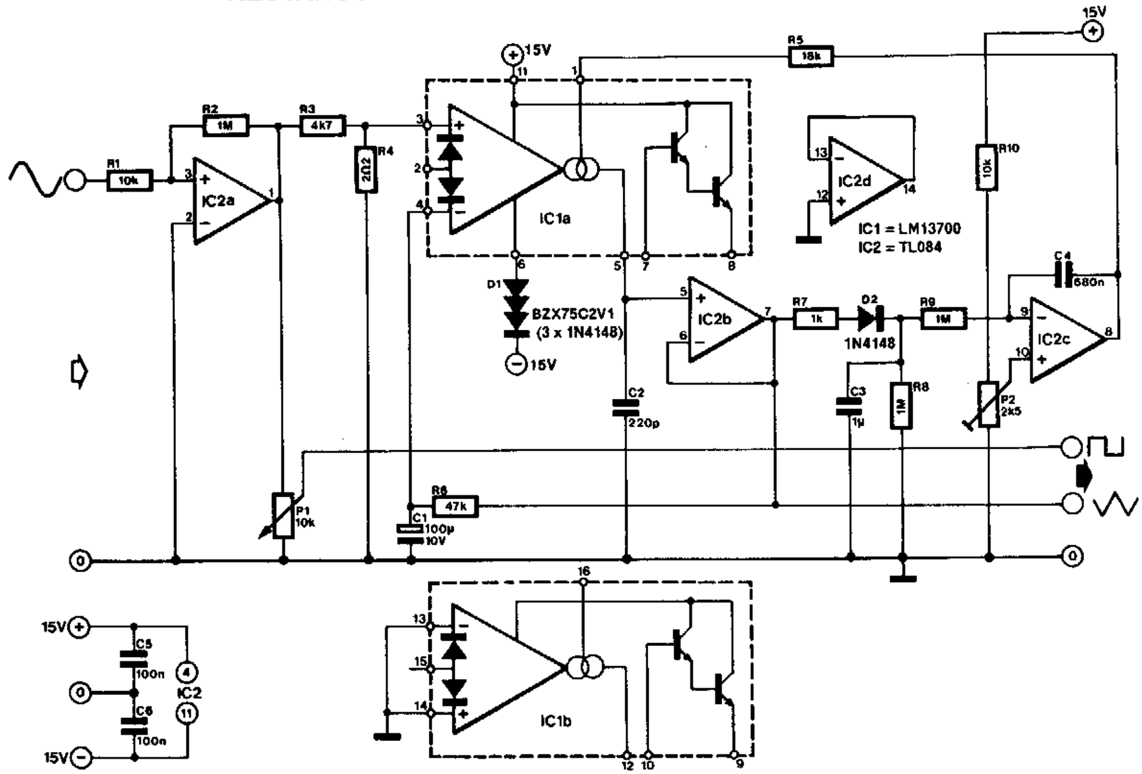
$$VC_{max} = \frac{I \times T}{C_1}$$

$$= \frac{6.4 \times t}{0.004\,7}$$

$$= 1362t$$

If $t = 1$ ms, then $VC_{max} = 1.362$ V.

When the input goes from low to high, a narrow positive pulse is generated at point A. This pulse becomes -13 V at point B, which cuts off D2. ICC's V+ voltage becomes zero. The charge on C1 will be absorbed by ICC on in a short time. The time constant of C2 and R5 determines the discharge period—about 10 μs. ICB is a buffer whose gain is equal to $(R_8 + R_9) \div R_9 = 1.545$. ICD's average voltage will be $(1362t \times 1.545) \div 2 = 1\,052t$. R10 and C3 smooth the sawtooth waveform to a dc output.



ELEKTOR ELECTRONICS *Fig. 21-6*

Many function generators are based on a rectangular waveform generator that consists of a Schmitt trigger and integrator. The triangular signal produced by the integrator is then used to form a sinusoidal signal with the aid of a diode network. The converter presented here works the other way around. It converts the output of a good-quality sine-wave oscillator into a rectangular and a triangular signal.

The sinusoidal signal is converted into a rectangular signal by IC2A. Because the output of this gate varies between −15 V and +15 V, it is reduced to a value that is suitable for integration by potential divider R3/R4. It is then integrated by transconductance amplifier IC1A and C2. The amplifier has a current output that is controlled by the current through pin 1. The output therefore behaves as a resistance, with which it is possible to influence the integration time.

The voltage across C2 is available in buffered form at the output of impedance inverter IC2B; this is the triangular signal. The amplitude of this signal is compared with a voltage set by P2 and the difference between these voltages, which is the output of IC2C, is applied to the current source at the output of IC1 via R5. This arrangement ensures that the level of the output voltage is virtually independent of the frequency of the rectangular signal or of the sinusoidal input.

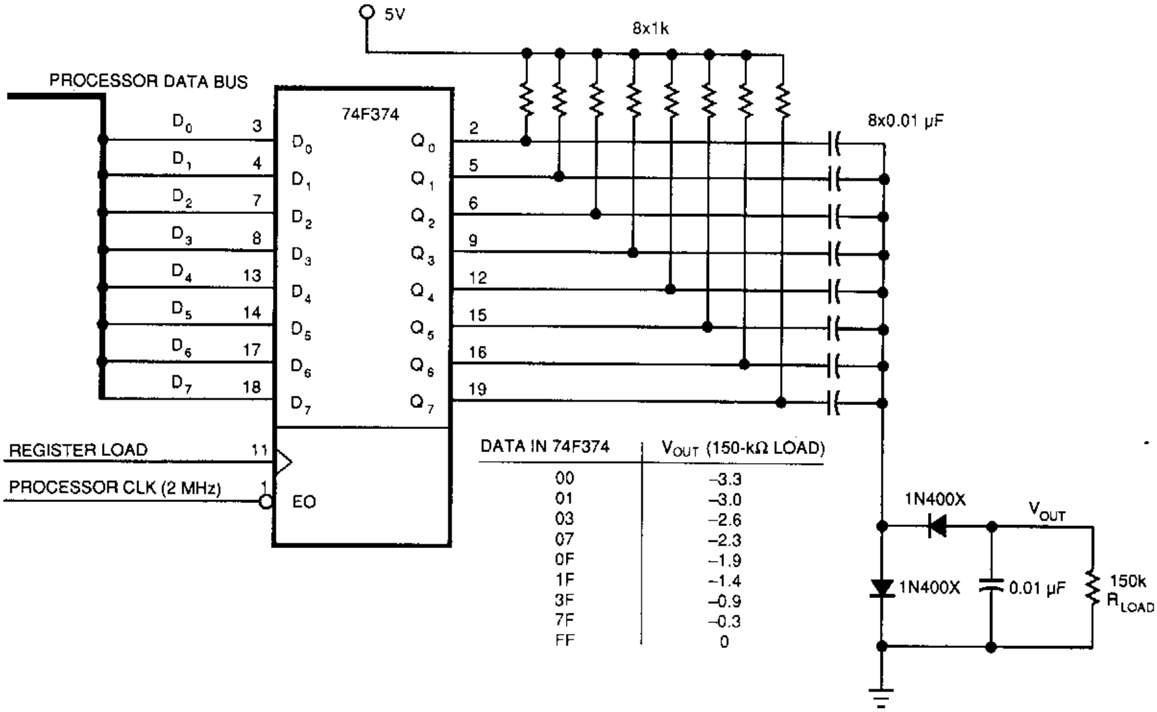
One problem with a precision integration is its being affected by offset voltages and bias currents. Feedback loop R6/C1 ensures that the output follows the potential across R4 accurately. However, tiny deviations might be caused by the bias current in circuit IC1, which is not greater than 8 μA at 70°C.

The time constant R6/C1 is large for a purpose: to ensure that the triangular signal, even at low frequencies, cannot affect the waveform of the signal to be integrated—the rectangular shape must be retained. The converter can process signals at frequencies from 6 Hz (where the amplitude is not affected) to 60 kHz (where the amplitude is reached by 10%).

Because of the long time constants, the time taken for the recovery of the amplitude of the triangular signal at frequencies above 1 kHz is rather long. The peak value of this signal should be set to 1 V.

Diode D1 is a so-called *stabistor*—three diodes in one package. It might be replaced by three discrete type 1N4148 diodes. The current drawn by the converter is of the order of 9 mA.

μP-CONTROLLED NEGATIVE VOLTAGE CONVERTER



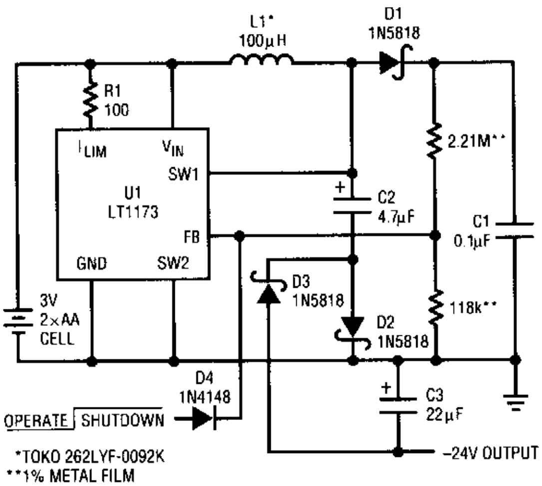
DATA IN 74F374	V_{OUT} (150-kΩ LOAD)
00	−3.3
01	−3.0
03	−2.6
07	−2.3
0F	−1.9
1F	−1.4
3F	−0.9
7F	−0.3
FF	0

EDN

Fig. 21-7

This circuit was used to produce a variable negative voltage for contrast control of an LCD display. A 74F374 generates a square wave that is ac coupled to a rectifier and load. By using the μP clock and data from the processor bus, and properly timed load signal, the dc level generated can be controlled by the μP.

dc/dc CONVERTER

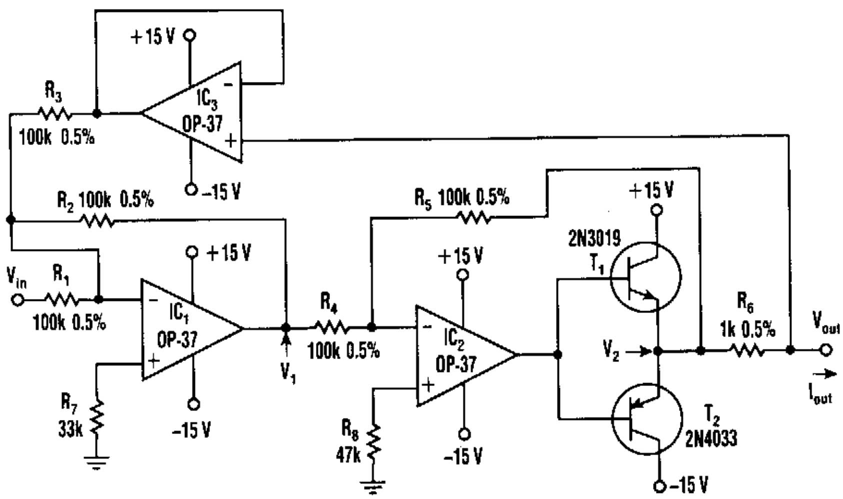


LINEAR TECHNOLOGY

Fig. 21-8

This circuit uses a Linear Technology LT1073 in a −24-V converter. The supply can be two AA cells (3 V) or 5 V. The circuit can deliver 7 mA.

VOLTAGE-TO-CURRENT CONVERTER



ELECTRONIC DESIGN

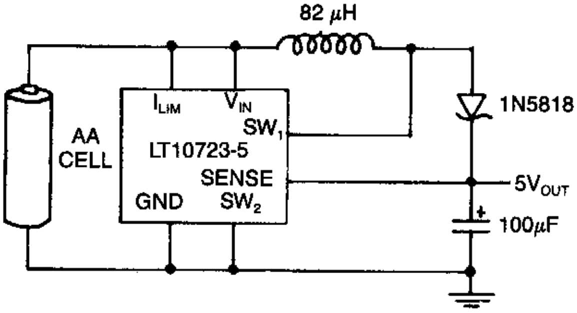
Fig. 21-9

This voltage to current converter uses three op amps to drive a pair of power transistors. The current output is calculated as:

$$I_{OUT} = \frac{V_{in}}{R_6}$$

Output resistance is over 50 MΩ. I_{OUT} can range from 1 mA to the current ratings of T1 and T2.

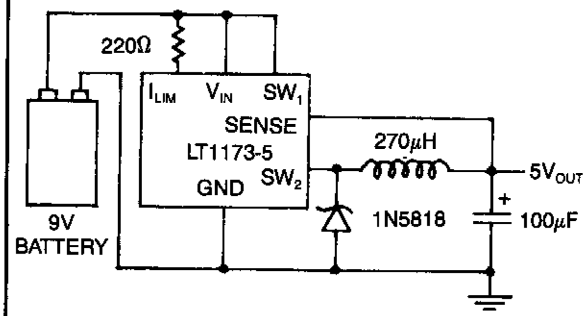
1-TO-5 V dc/dc CONVERTER



LINEAR TECHNOLOGY *Fig. 21-10*

This circuit, using the Linear Technology LT1173, produces 5 V at 40 mA from a 1.5-V AA cell.

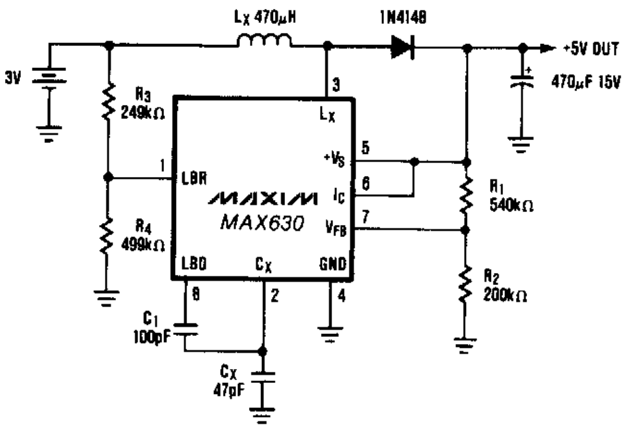
9-TO-5-V CONVERTER



LINEAR TECHNOLOGY *Fig. 21-11*

Using a Linear Technology LT1173-5, this converter produces regulated 5 V from a 9-V battery.

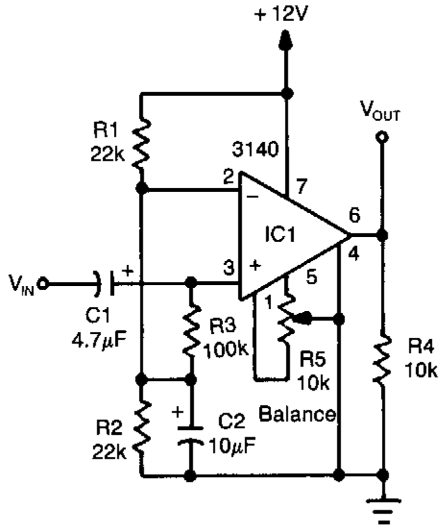
+3-V BATTERY TO +5-V dc/dc CONVERTER



MAXIM *Fig. 21-12*

A common power-supply requirement involves converting a 2.4- or 3-V battery voltage to a 5-V logic supply. This circuit converts 3 V to 5 V at 40 mA with 85% efficiency. When I_C (pin 6) is driven low, the output voltage will be the battery voltage minus the drop across diode D1. The optional circuitry that uses C1, R3, and R4 lowers the oscillator frequency when the battery voltage falls to 2.0 V. This lower frequency maintains the output-power capability of the circuit by increasing the peak inductor current, which compensates for the reduced battery voltage.

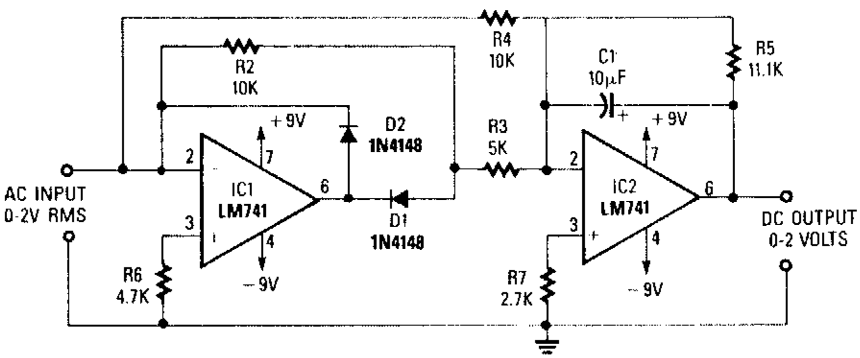
SINE-WAVE/SQUARE-WAVE CONVERTER



RADIO-ELECTRONICS *Fig. 21-13*

An op amp used as a comparator produces a 10-V p-p square-wave output with 100-mV input, to 15 kHz. Adjust R5 for symmetry of square wave at low input levels.

PRECISION FULL-WAVE ac/dc CONVERTER



RADIO-ELECTRONICS *Fig. 21-14*

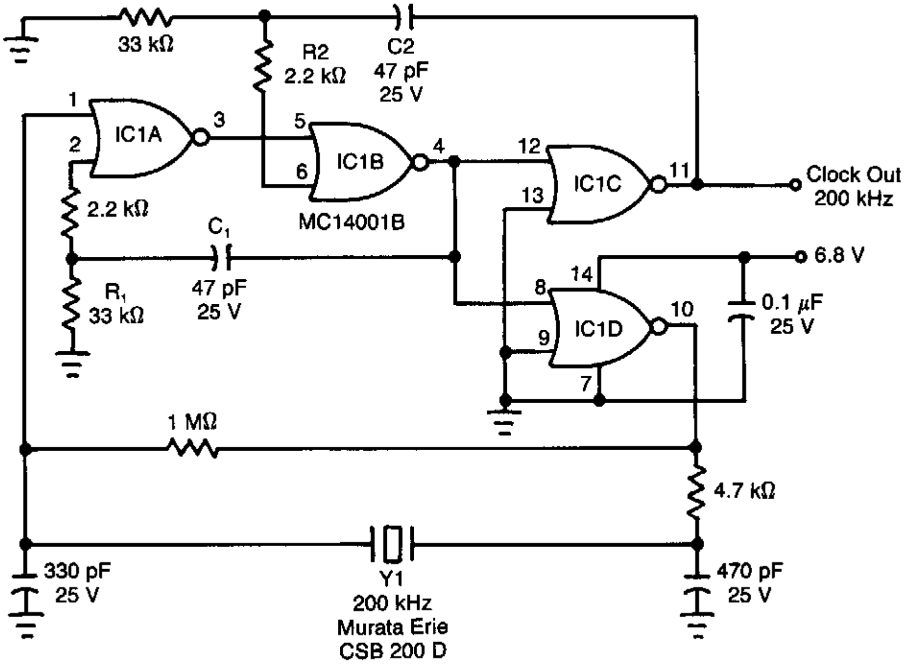
A dc level is produced that corresponds to the ac input rms value (if sine wave). $\frac{R_1}{C_5}$ set the gain of IC2 to 1.11. This factor is the average-to-rms conversion factor. IC1 and IC2 act as a full-wave rectifier circuit, with D1 and D2.

22

Crystal Oscillators

The sources of the following circuits are contained in the Sources section, which begins on page 664. The figure number in the box of each circuit correlates to the entry in the Sources section.

MICROPOWER CLOCK



EDN

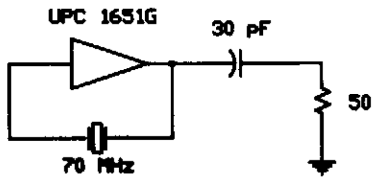
Fig. 22-1

Although ceramic resonators are a good choice for low-power, low-frequency clock sources (if you can stand their 30-ppm temperature coefficient), they have troublesome, spurious-resonance modes. This circuit rejects all but the resonator's fundamental mode. This clock circuit works from −40 to +80°C and consumes only 2.8 mW.

The rising edge of resonator Y1 toggles IC1A low. ac-coupled positive feedback from IC1D via C1 and R1 immediately confirms this state change at IC1B so that Miller loading, harmonic components, or below-minimum rise times at IC1A cannot force IC1C to relapse to its previous state. This tactic also applies to resonator Y1's falling edge because IC1C, via C2 and R2, holds IC1B high.

Choose time constants R_1C_1, and R_2C_2 to be equal and ranging from 60 to 75% of one-half of the clock's period. Ceramic capacitors (10% tolerance) with X7R dielectric work well. With these time constants, the logic will be locked and unavailable to the ceramic resonator until just before it executes a legitimate transition. IC1D and IC1C are in parallel to isolate the resonator from external loads and, more importantly, from C2.

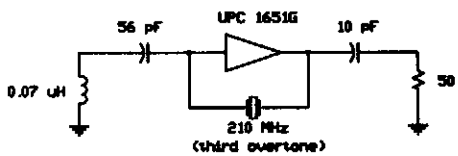
SIMPLE FUNDAMENTAL CRYSTAL OSCILLATOR



RF DESIGN Fig. 22-2

This simple fundamental oscillator uses a μPC1651G IC and two components. The crystal is fundamental.

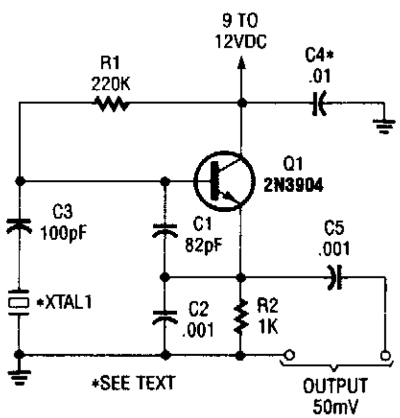
SIMPLE THIRD-OVERTONE OSCILLATOR



RF DESIGN Fig. 22-3

Using a 210-MHz third overtone crystal, this circuit operates directly at the crystal frequency, with 210-MHz output and no multiplier stages.

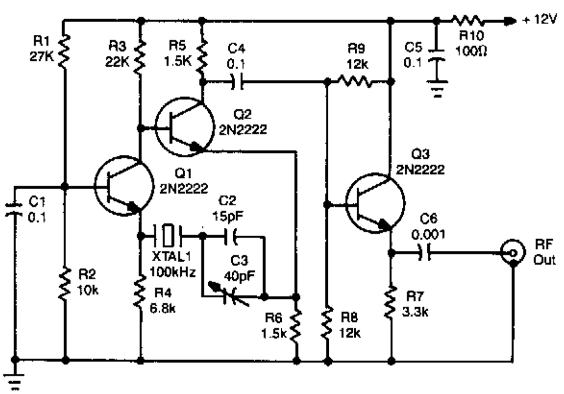
COLPITTS 1-to 20-MHz CRYSTAL OSCILLATOR



This is a simple Colpitts crystal oscillator for 1 to 20 MHz, can be easily made from junk-box parts (provided that a crystal is handy).

POPULAR ELECTRONICS Fig. 22-4

100-kHz CRYSTAL CALIBRATOR

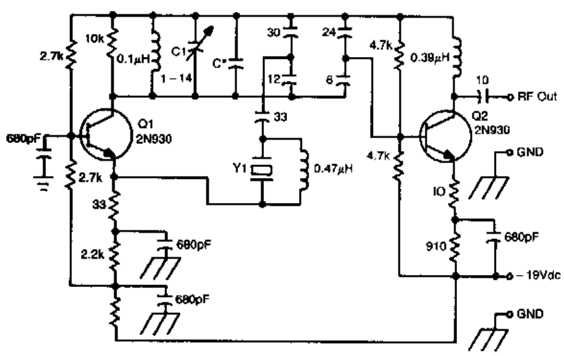


POPULAR ELECTRONICS

Fig. 22-5

Using a 12-V supply, this crystal calibrator should prove a useful accessory for a SW receiver. Q1 and Q2 form an oscillator and Q3 is a buffer amp.

100-MHz OVERTONE OSCILLATOR

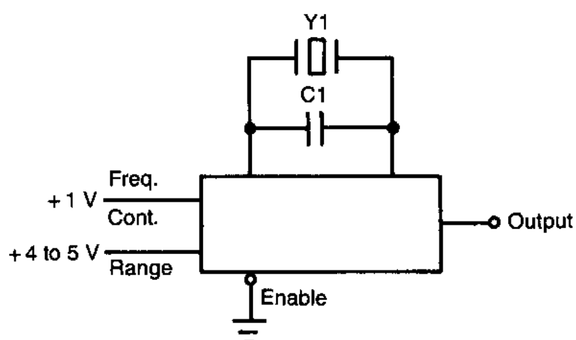


73 AMATEUR RADIO

Fig. 22-6

This oscillator circuit uses a 5th overtone crystal in the 85-to-106 MHz range. Y1 is the crystal. The circuit was originally used to frequency control a microwave oscillator.

VOLTAGE-CONTROLLED CRYSTAL OSCILLATOR



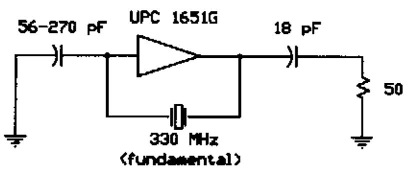
Notes:

1. For frequencies ≤ 1 MHz, C1 = 5 to 15 pF. For frequencies ≥ 1 MHz, C1 can be eliminated.
2. IC is SN74S124 for f_{max} of 60 MHz. IC is SN74LS124 for f_{max} of 35 MHz.

VALPEY-FISHER

Fig. 22-7

330-MHz CRYSTAL OSCILLATOR

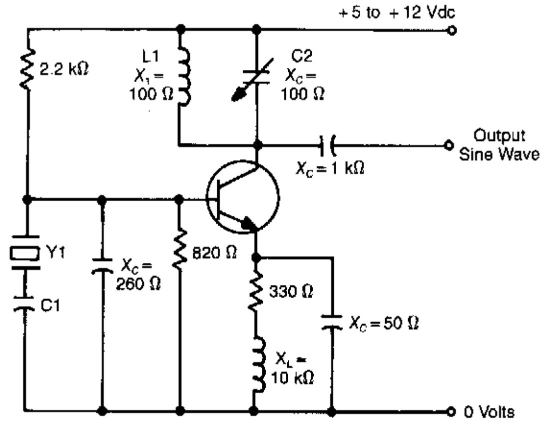


RF DESIGN

Fig. 22-8

A μPC 1651G IC operates in the fundamental mode with an experimental crystal at 330 MHz. The 56-to-270 pF capacitor is not critical; about +1-dBm RF output is available.

10-to 150-kHz OSCILLATOR



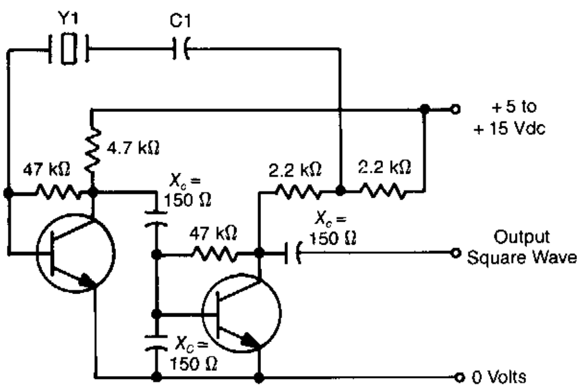
VALPEY-FISHER

Fig. 22-10

Note:

Y1 is "H", "NT," or "E" cut

10-to 80-MHz OSCILLATOR



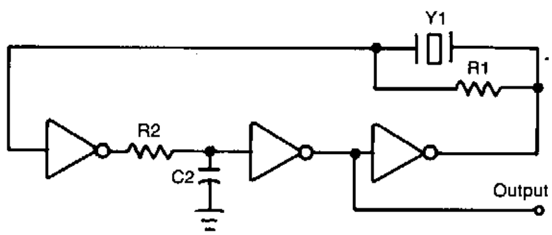
VALPEY-FISHER

Fig. 22-9

Notes:

1. Y1 is "AT" cut, fundamental, or overtone crystal.
2. Tune L1 and C2 to operating frequency.

1-to 4-MHz CMOS OSCILLATOR



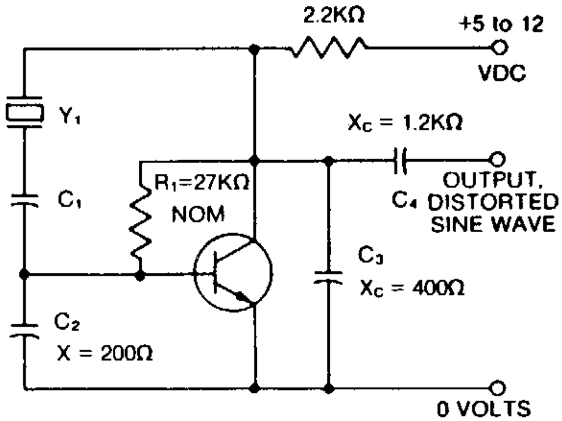
VALPEY-FISHER

Fig. 22-11

Notes:

1. 1 M$\Omega < R_1 <$ 5 MΩ
2. Select R_2 and C_2 to prevent spurious frequency.
3. ICs are 74C04 or equivalent.

150-to 30 000-kHz OSCILLATOR



Notes:

1. Y1 is "AT", "CT", "DT", "NT", "SL", or "E" cut.
2. C1, C2, and C3 in series should equal the load capacity of crystal.
3. Adjust R1 for 1/2 supply voltage at collector of transistor.

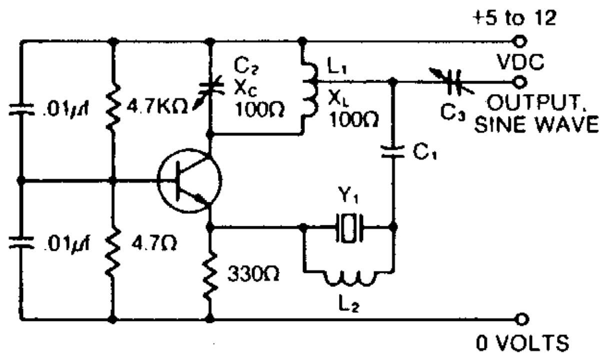
VALPEY-FISHER

Fig. 22-12

C1 capacitor in series with the crystal may be used to adjust the output frequency of the oscillator. The value can range between 20 pF and 0.01 μF, or it can be a trimmer capacitor.

X values are approximate and can vary for most circuits and frequencies; this is also true for resistance values. Adequate power supply decoupling is required; local decoupling capacitors near the oscillator are recommended.

50-to 150-MHz OSCILLATOR



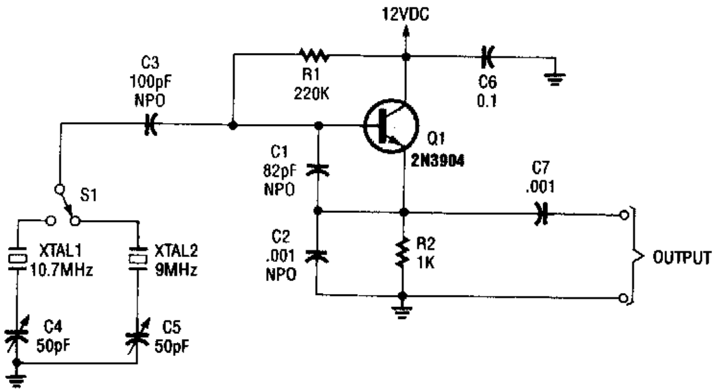
VALPEY-FISHER

Fig. 22-13

Notes:

1. Y1 is "AT" cut, overtone crystal.
2. Tune L1 and C2 to operating frequency.
3. L2 and shunt capacitance (C0) of crystal (approximately 6 pF) should resonate to the output frequency of the oscillator (L2 = 0.5μH at 90 MHz). This is necessary in order to tune out effect of C0 of the crystal.
4. C3 is varied to match output.

TWO-FREQUENCY COLPITTS OSCILLATOR

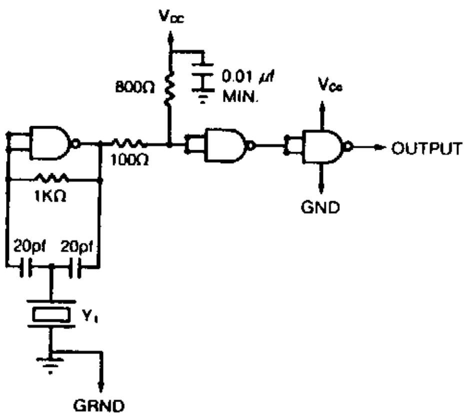


POPULAR ELECTRONICS

Fig. 22-14

Using switched crystals, this oscillator is intended for receiver alignment purposes.

1-to 20-MHz TTL OSCILLATOR



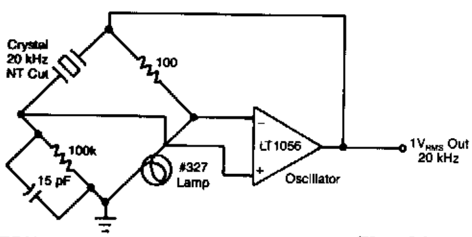
Notes:

1. Y1 is "AT" cut fundamental crystal.
2. ICs are 7400/7404.

VALPEY-FISHER

Fig. 22-15

CRYSTAL-CONTROLLED BRIDGE OSCILLATOR

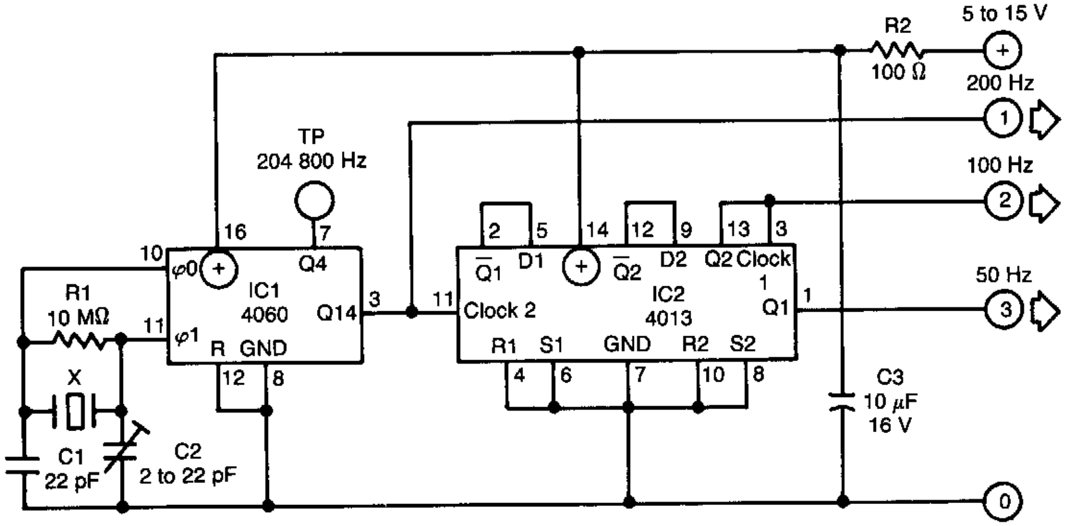


This crystal-controlled oscillator uses the current variations in a small lamp to stabilize amplitude variations.

EDN

Fig. 22-16

ECONOMICAL CRYSTAL TIME BASE

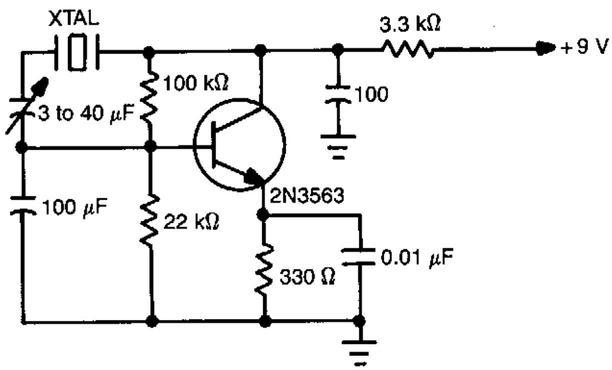


ELEKTOR ELECTRONICS

Fig. 22-17

The above time base circuit will provide 50-, 100-, or 200-Hz signals from an inexpensive crystal cut for 3.276 8 MHz, a common crystal used for microprocesser works. It requires a power supply of 5 to 15 V at 0.05 to 2.5 mA.

CRYSTAL OSCILLATOR



WILLIAM SHEETS

Fig. 22-18

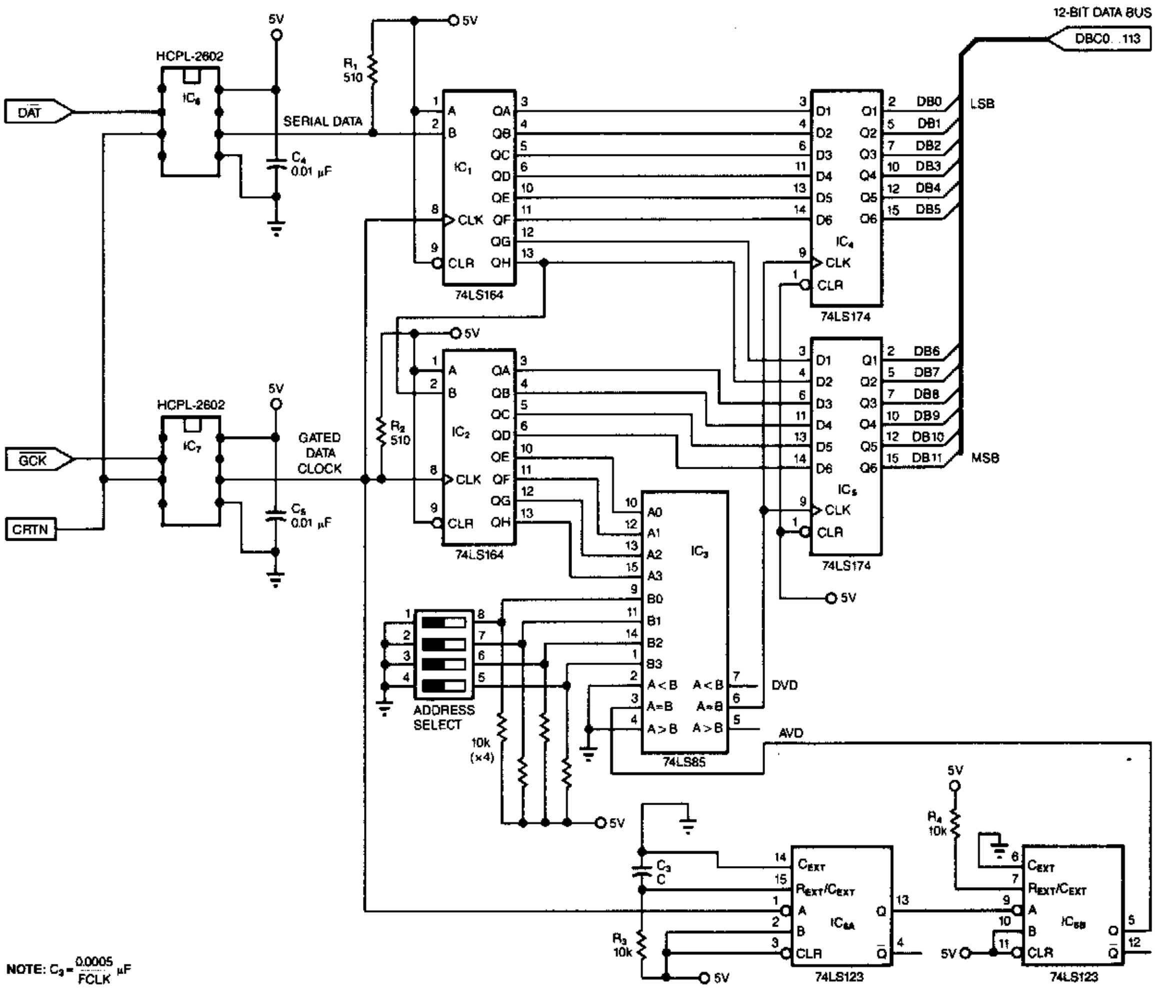
This simple circuit will oscillate with a wide range of crystals. Connect several different types of crystal holders in parallel to improve versatility. The 3-to 40-pF capacitor adjusts crystal frequency over a small range for setting to standard-frequency transmissions when the unit is used as a crystal calibrator.

23

Data Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 664. The figure number in the box of each circuit correlates to the entry in the Sources section.

3-WIRE RECEIVER/MESSAGE DEMUXER

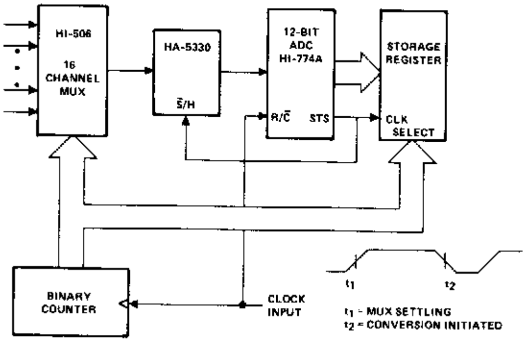


EDN

Fig. 23-1

This 3-wire receiver checks the first four data bits of 16 received bits against a preset address. If the two match, the remaining 12 bits are latched into two 6-bit flip-flop registers. Either CMOS or TTL logic families can be used in the design.

DATA ACQUISITION SYSTEM I

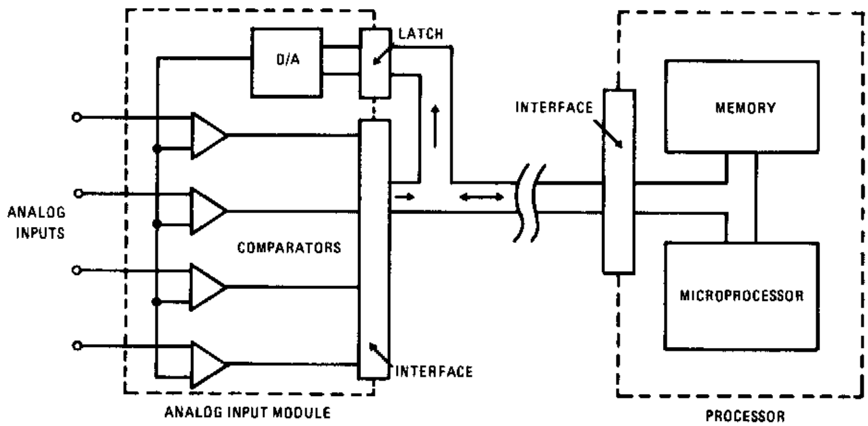


HARRIS *Fig. 23-2*

The HI-506 multiplexer is used as an analog input selector, controlled by a binary counter to address the appropriate channel. The HA-5330 is a high-speed sample and hold. The sample/hold control is tied to the status (STS) output of the HI-774A; whenever a conversion is in process, the S/H is in the hold mode. A conversion is initiated when the clock input becomes low; when the clock becomes high, the mux address changes. The mux will be acquiring the next channel while the ADC is converting the present input, held by the S/H. The clock low time should be between 225 ns and 6.5 μs, with the period greater than 8.5 μs.

With this timing, T/C will be high at the end of a conversion, so the output data will be valid ~ 100 ns before STS goes low. This allows STS to clock the data into the storage register. The register address will be offset by one; if this is a problem, a 4-bit latch can be added to the input of the storage register. With a 100-kHz clock rate, each channel will be read every 160 μs.

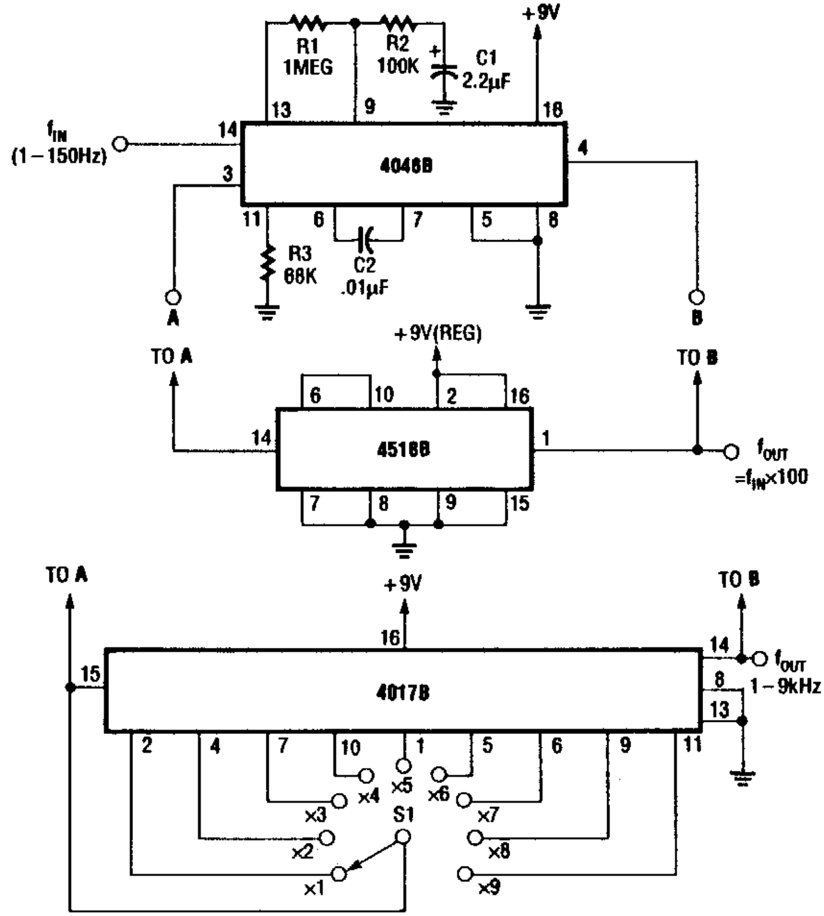
DATA ACQUISITION SYSTEM II



HARRIS *Fig. 23-3*

In this circuit, an HA-4900 series comparator is used in conjunction with a D/A converter to form a simple, versatile, multichannel analog input for a data acquisition system. The processor first sends an address to the D/A, then the processor reads the digital word generated by the comparator outputs. To perform a simple comparison, the processor sets the D/A to a given reference level, then examines one or more comparator outputs to determine if the inputs are above or below the reference. A window comparison consists of two such cycles with two reference levels set by the D/A. One way to digitize the inputs would be for the processor to increment the D/A in steps. The D/A address, as each comparator switches, is the digitized level of the input. While stairstepping, the D/A is slower than successive approximation; all channels are digitized during one staircase ramp.

LOW-FREQUENCY PRESCALER



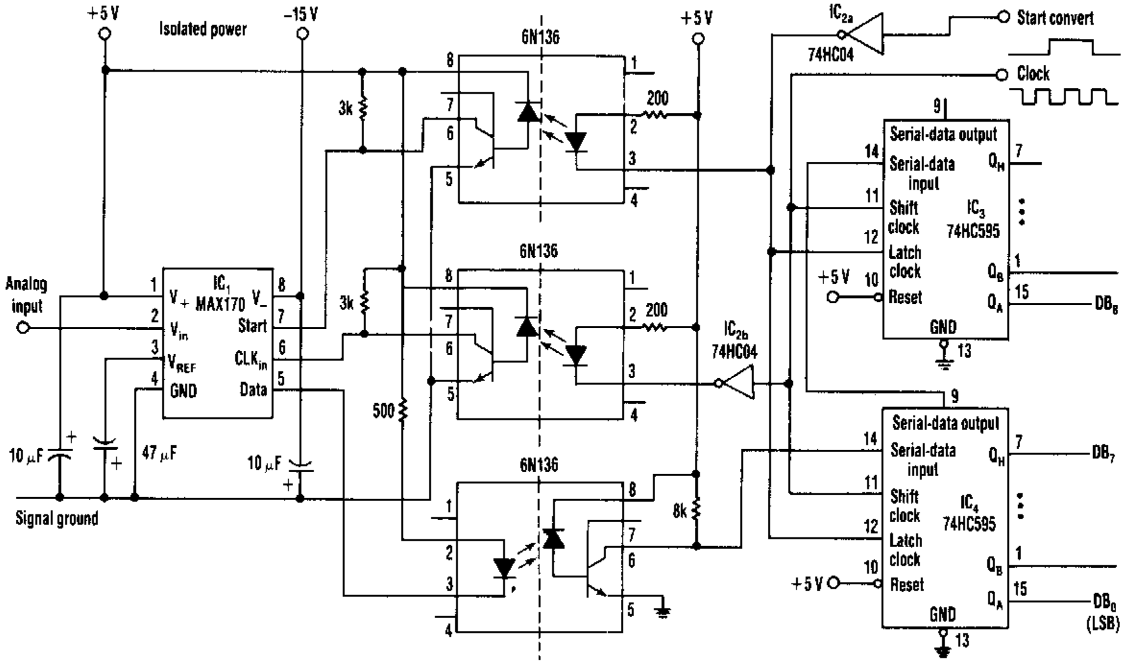
RADIO-ELECTRONICS

Fig. 23-4

For multiplying frequencies in the 1-to 150-Hz range, this circuit uses a 4046B and a ÷100 prescaler. The VCO output is phaselocked to the low-frequency input. This simplifies use of a frequency counter to measure LF signal frequencies.

By using a 4017B and a 1-kHz f_{IN}, the circuit can be used as a 1-to 9-kHz frequency synthesizer or as a ×10 frequency multiplier.

ANALOG DATA-SIGNALS ISOLATER



ELECTRONIC DESIGN

Fig. 23-5

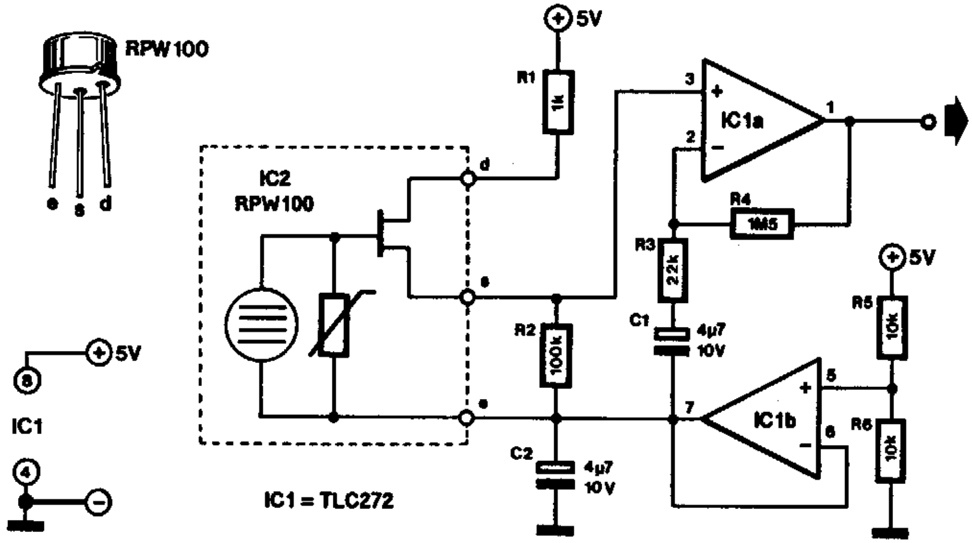
By converting analog data to digital and using optocouplers, this circuit can be used to transmit analog signals across barriers, such as voltage levels, different ground systems, etc.

24

Detectors

The sources of the following circuits are contained in the Sources section, which begins on page 664. The figure number in the box of each circuit correlates to the entry in the Sources section.

THERMALLY OPERATED DIRECTION DETECTOR



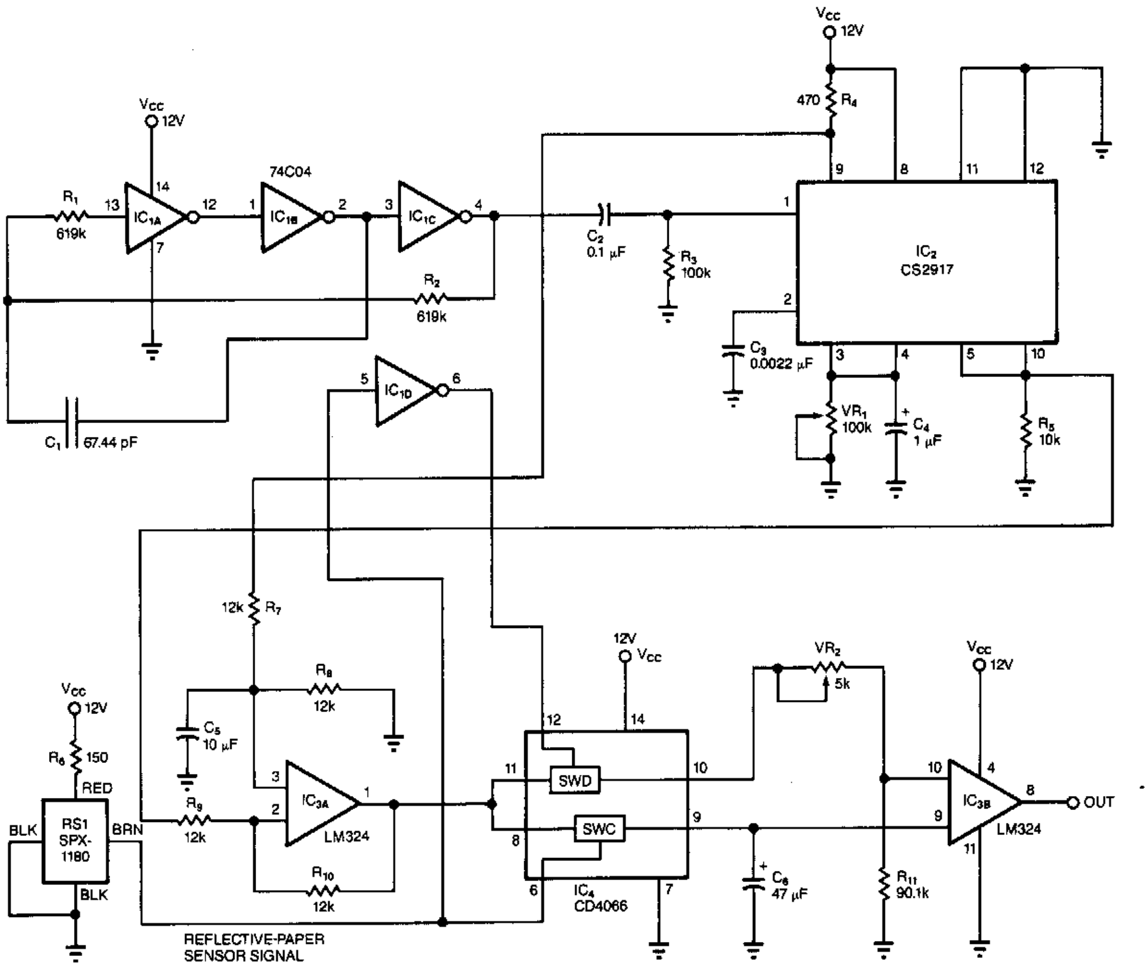
ELEKTOR ELECTRONICS *Fig. 24-1*

A heat-sensitive sensor can be used to construct a direction detector. Such a sensor reacts to all animal heat. The one used in this design has a sensitive surface that has been divided into two. It, therefore, makes a difference, whether the heat approaches from the left or the right. The indication for cold objects is, of course, exactly the opposite.

Circuit IC1B forms a symmetric supply. Terminal s of the sensor is its output. The signal at s is amplified in IC1A by a factor of about 70 before it is available at the output of the detector.

To obtain good directivity, it is best to place the sensor behind a single narrow slit, rather than behind the usual raster of a multifacetted mirror. The circuit draws a current of only a few mA from a 5-V supply.

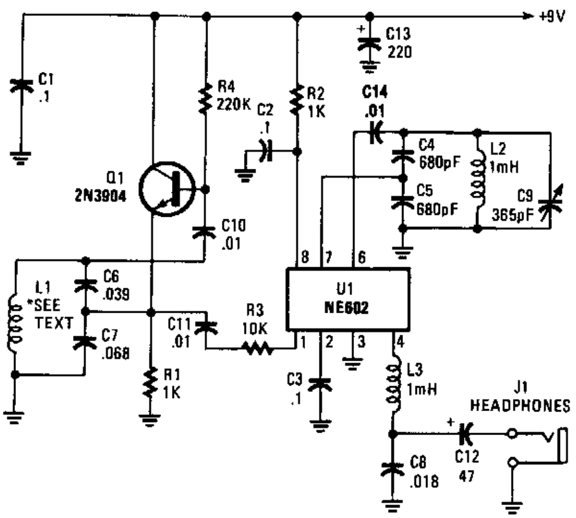
TWO-SHEETS DETECTOR



EDN *Fig. 24-2*

Using the principle of capacitance between two plates this circuit senses when more than one sheet of paper goes between the sensing electrodes at a time. C1 is the sensing capacitor formed of two plates. It consists of two plates 2″ × 15″ with 0.1″ spacing. A change of capacitance causes a change in oscillator frequency of the IC1 circuit, which is detected by IC2 and IC3.

METAL DETECTOR

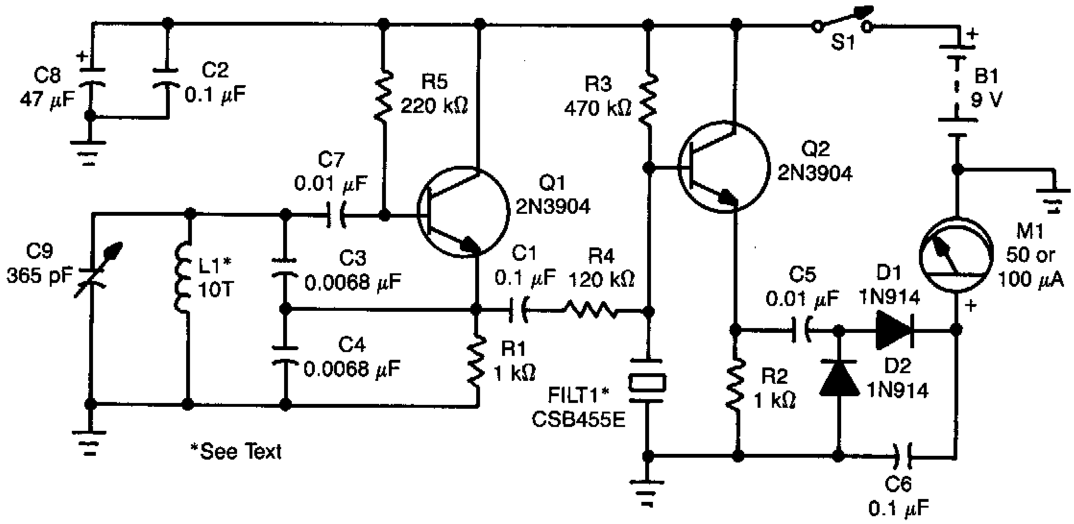


An NE602 acts as a heterodyne detector and Q1 as a sense oscillator. When L1 is brought near metal, it causes a charge in loop inductance, shifting the resonant frequency of L1 C6/C7. L1 is 5 turns #20 wire on a 9″ diameter wood or plastic form.

POPULAR ELECTRONICS

Fig. 24-3

METAL DETECTOR

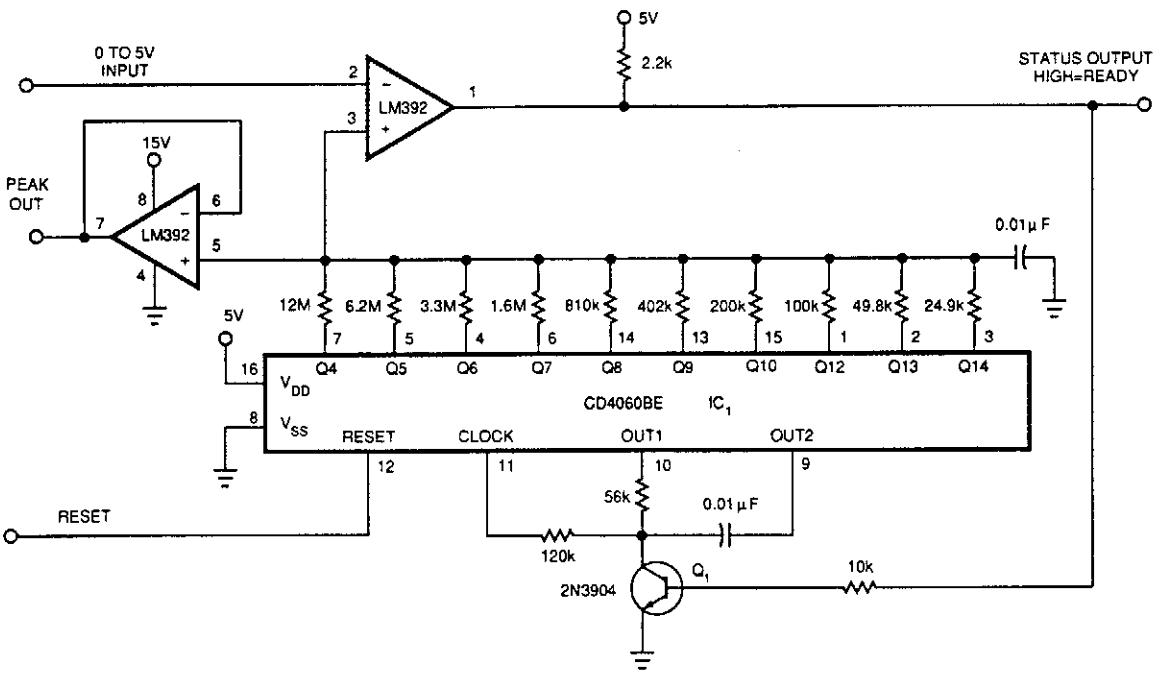


POPULAR ELECTRONICS

Fig. 24-4

Using an oscillator running at 455 kHz, the metal-detector circuit produces an indication on the meter M1. When the oscillator frequency changes because of metal in the field of L1, the change will show as an increase or decrease in frequency, which produces a charge in the meter reading. The ceramic filter FILT1 produces a selective bandpass that yields this effect. L1 can be a 4″ diameter coil wound on a suitable plastic form. About 10 turns of #26 wire are required. Use a frequency counter to adjust L1 and verify that Q1 is operating on or near 455 kHz.

PEAK DETECTOR

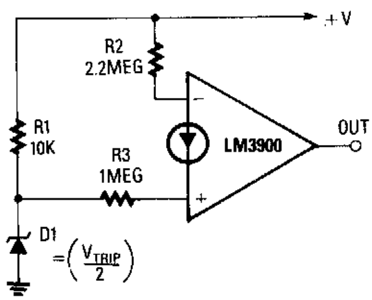


EDN

Fig. 24-5

A 0-to-5 V input drives the negative input of LM392 comparator if reset (pin 12) if DC4060BE is pulled high then low, all outputs of ICF1 are forced low, forcing + input of comparator to go low. Q1 is cut off and IC1's clock oscillator, running at about 775 Hz, starts counting. The Q4 through Q14 outputs connect to a ladder. When the counter reaches a count so that the voltage on pin 3 of the LM392 equals the peak input voltage, the counter stops. This voltage is available at the output of the voltage follower LM392 (pin 7). The maximum time to acquire a peak is 22 seconds. This circuit is slow and was originally intended for battery-charging applications.

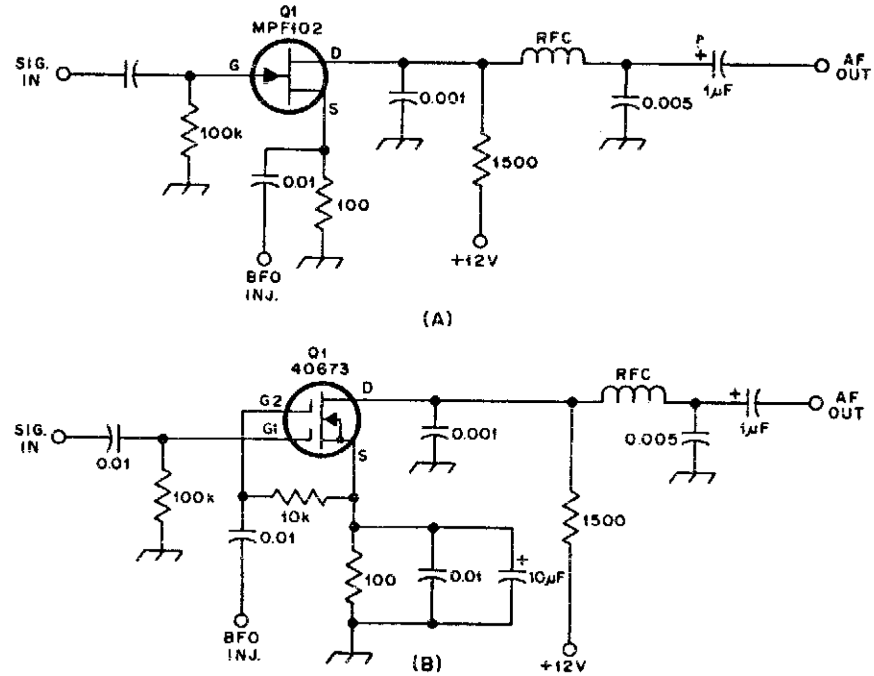
UNDERVOLTAGE DETECTOR



RADIO-ELECTRONICS *Fig. 24-6*

The output goes high when the supply falls below a value determined by zener diode D1. If D1 is a 5.6-V zener, the op amp will switch high when the supply voltage falls below approximately 11 V. The precise trip point can be varied by replacing R3 with an 820-kΩ resistor in series with a 470-kΩ potentiometer.

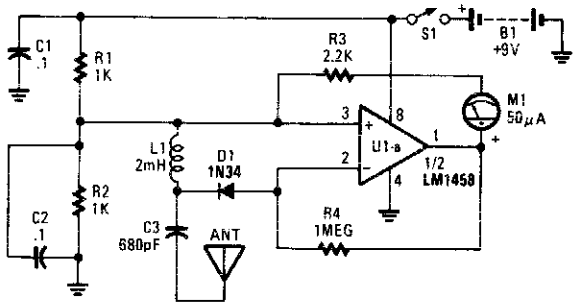
SSB/CW PRODUCT DETECTORS



ARRL HANDBOOK *Fig. 24-7*

These circuits are used for product detection of single-sidebound (SSB) and CW signals. BFO injection is typically 0.5 to 1 V rms for both circuits. Frequencies can be up to 25 MHz or so.

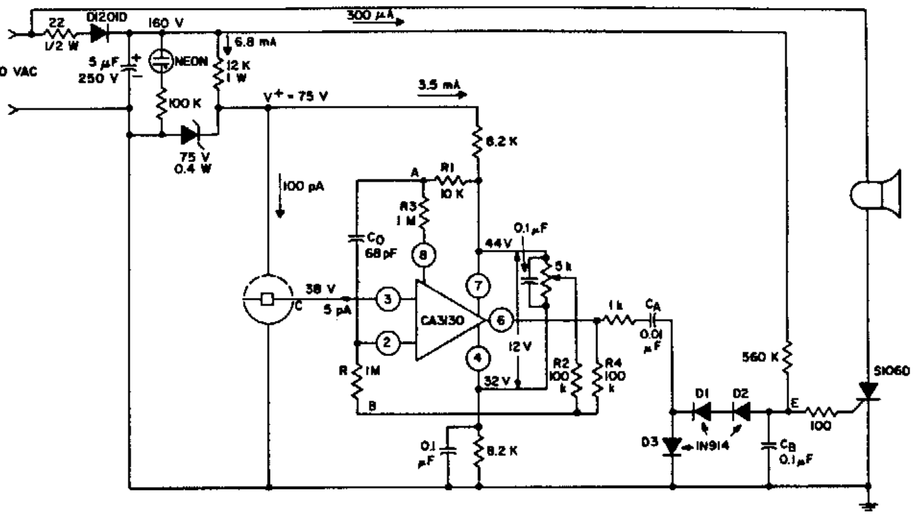
RF-FIELD DETECTOR



POPULAR ELECTRONICS *Fig. 24-8*

This detector is a half-wave rectifier for RF, which then feeds an op amp. U1A acts as an amplifier, driving meter M1. This circuit can detect mW RF levels from below the AM broadcast band to well above the FM broadcast band.

LINE-OPERATED SMOKE DETECTOR

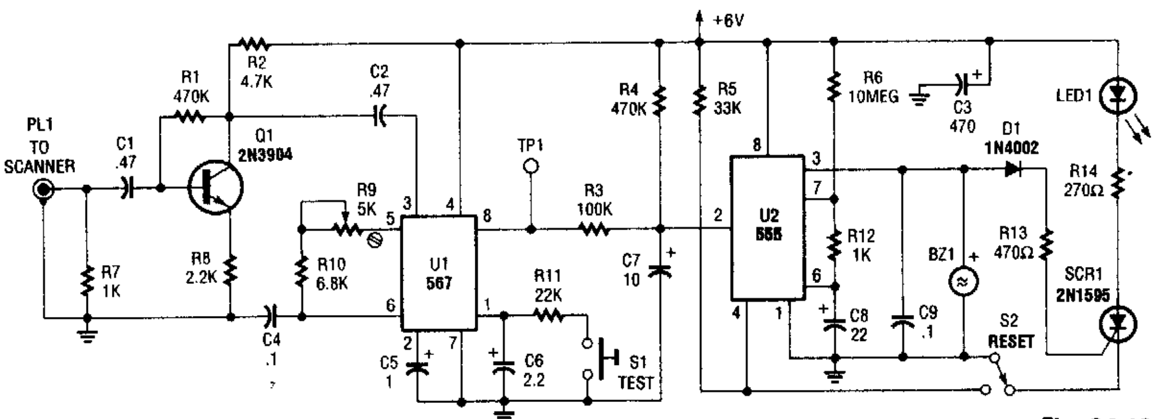


RCA

Fig. 24-9

Using an ionization chamber and a high-impedance (CA3130) op amp, the presence of smoke will cause the CA3130 to stop oscillating, triggering S106D SCR, sounding the alarm.

NOAA WEATHER ALERT DECODER



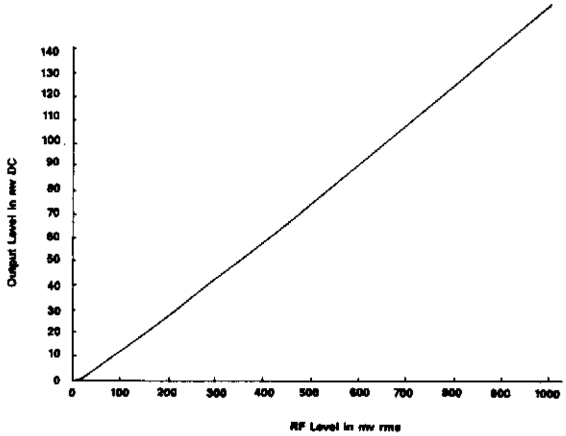
POPULAR ELECTRONICS

Fig. 24-10

This circuit detects the 1050-Hz tone sent by the NOAA (National Oceanic and Atmospheric Administration) Weather radio stations that operate from 162.40 to 162.55 MHz. This tone lasts for several seconds. Q1 is an amplifier that feeds tone detector U1, an NE567 detects this tone and produces a low on pin 8. This is coupled to a 555 timer (U3), which produces a high on its pin 3, sounds BZ1, triggers SCR1, and lights the LED. S2 is used to rest the circuit.

Audio is taken from the receiver that is used with the device. S1 is used to test the device and it will sound the alarm in two seconds if all is OK.

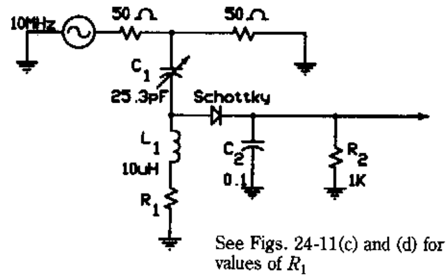
LOW-LEVEL DIODE ENVELOPE DETECTOR



Envelope detector response.

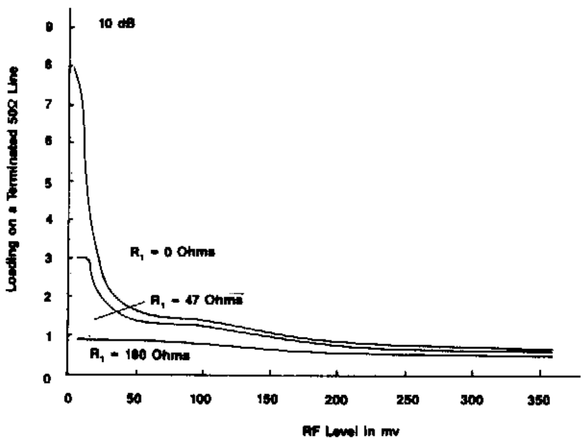
RF DESIGN

Fig. 24-11(a)



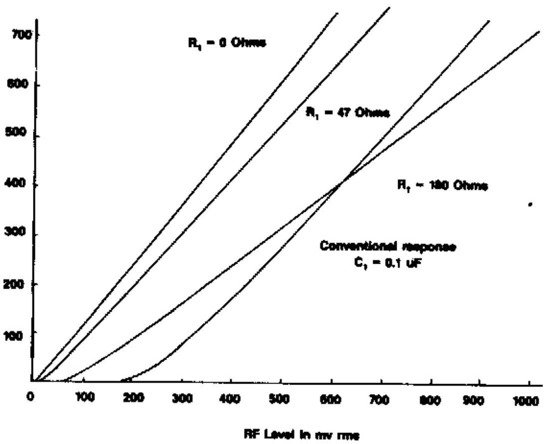
Low-level envelope detector with lower L/C ratio to illustrate the effect of R.

Fig. 24-11(c)



Terminated 50-ohm line.

Fig. 24-11(b)

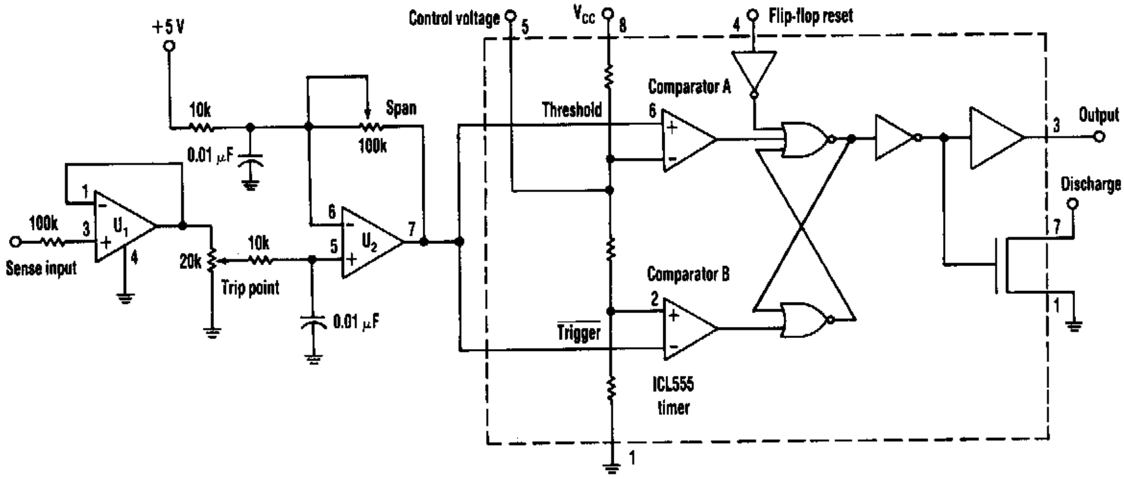


Envelope detector response for 3 values of R.

Fig. 24-11(d)

An approach to low-level RF detection and performance curves is shown here. This design is for 10 MHz, but values can be scaled to other frequencies, if needed.

DETECTOR AND CONTROLLER



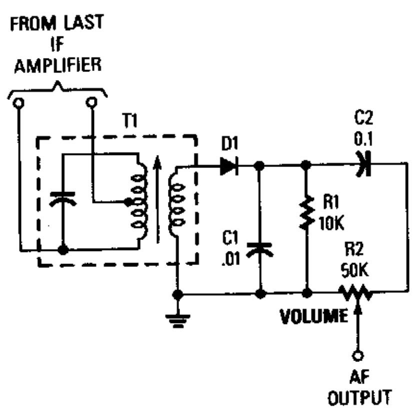
ELECTRONIC DESIGN

Fig. 24-12

Many applications require analog signals to be sensed and digital signals to be controlled. A way to detect these points is by using a 555 timer in an unconventional configuration. This method will also add hysteresis to the circuit and guard against oscillation. The 555 supplies two comparators and a flip-flop eliminates the oscillation. Using this classic timer in the new configuration also reduces the component count.

The circuit shows the 555's trigger and threshold pins tied together. This enables the comparators to set and reset the flip-flop. Op amp U2 supplies both the trip-point setting and a way to adjust the hysteresis for ON and OFF points. One application where this circuit would be useful is in a Nicad battery-charge controller.

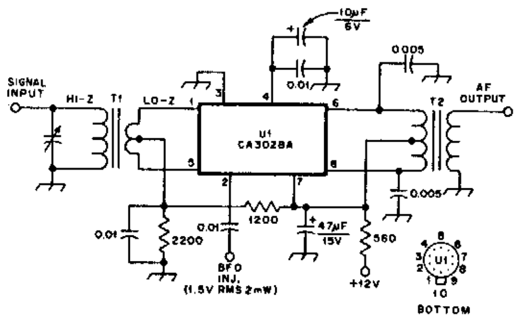
AM ENVELOPE DETECTOR



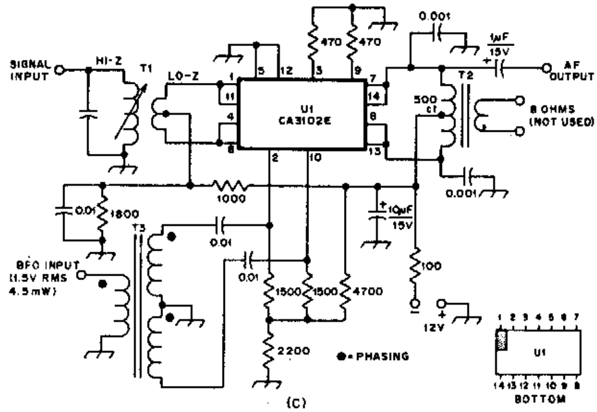
POPULAR ELECTRONICS *Fig. 24-13*

This general-purpose detector for AM envelope detection can be used in many receiver applications. T1 matches the IF amplifier impedance (typically 1 to 10 kΩ) to the 1 kΩ (approximately) detector impedance. D1 can be an IN60, IN82AG, IN270, or a similar type.

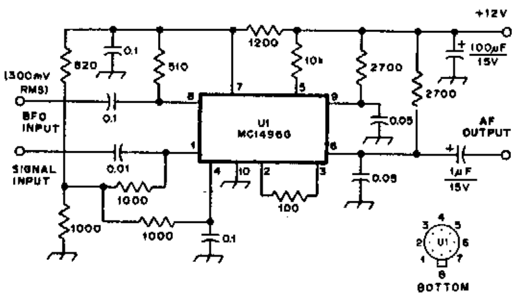
IC PRODUCT DETECTORS



ARRL HANDBOOK *Fig. 24-14(a)*



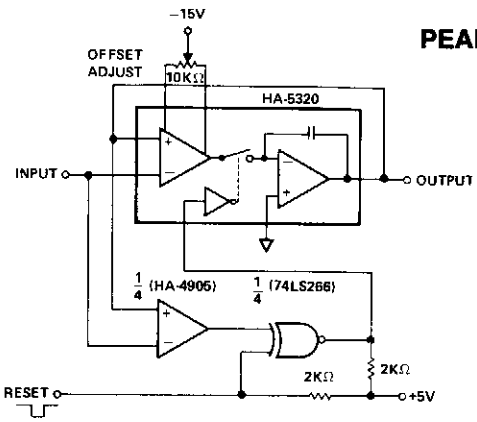
ARRL HANDBOOK *Fig. 24-14(c)*



ARRL HANDBOOK *Fig. 24-14(b)*

These product detectors use IC devices. SSB and CW signals can be detected with them. The circuits should be useful up to 20 or 30 MHz. T3 in (c) is a 1:1:1 toroidal type, depending on the BFO frequency.

PEAK DETECTOR

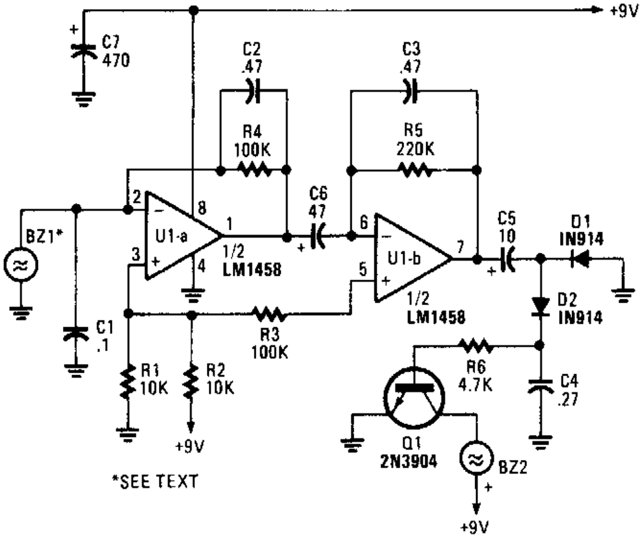


HARRIS *Fig. 24-15*

An analog signal requires about 100 ns to propagate through the HA-5320. For time-varying signals, this assures a voltage difference between input and output. Also, the voltage changes polarity when the signal slope changes polarity (passes a peak). This behavior makes the circuit a possible sample/hold peak detector, by adding a comparator to detect the polarity changes.

The exclusive NOR gate allows a reset function which forces the HA-5320 to the sample mode. The connections shown detect positive peaks; the comparator inputs can be reversed to detect negative peaks. Also, the offset must be introduced to provide enough step in voltage to trip the comparator after passing a peak. This circuit works well from below 100 Hz up to the frequency at which slew-rate limiting occurs. It captures the amplitude of voltage pulses, provided that the pulse duration is sufficient to slew to the top of the pulse.

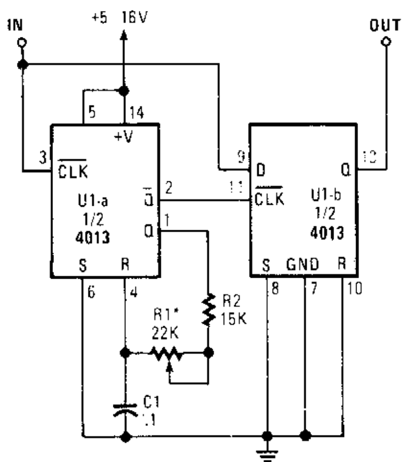
AIR-PRESSURE CHANGE DETECTOR



POPULAR ELECTRONICS *Fig. 24-16*

A piezoelectric detector (BZ1) is used in this circuit to detect a change in air pressure. BZ1 produces a voltage that is amplified by U1A and U1B. Frequency response is limited to low frequencies. The signal is rectified by D1 and D2 and drives Q1, which activates BZ2, a piezoelectric buzzer.

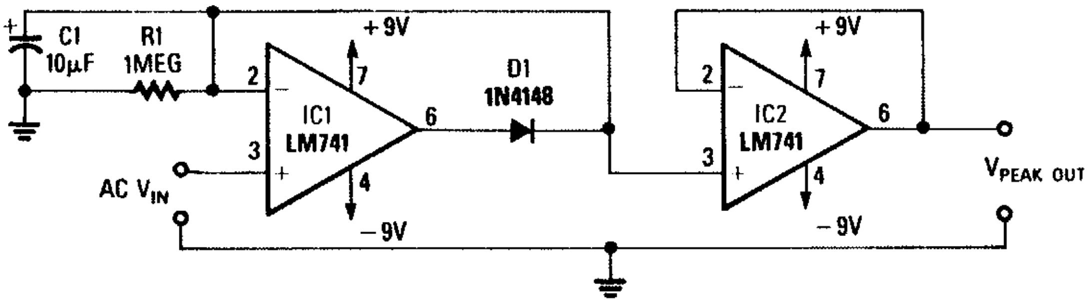
DUTY-CYCLE DETECTOR



POPULAR ELECTRONICS *Fig. 24-17*

This circuit looks at the time an incoming pulse is high. If the incoming pulse is shorter than the adjusted (VAR1) pulse, the output of U1B is high. Values are shown for a 1-to 2-μs pulse.

OP AMP PEAK DETECTOR



RADIO-ELECTRONICS *Fig. 24-18*

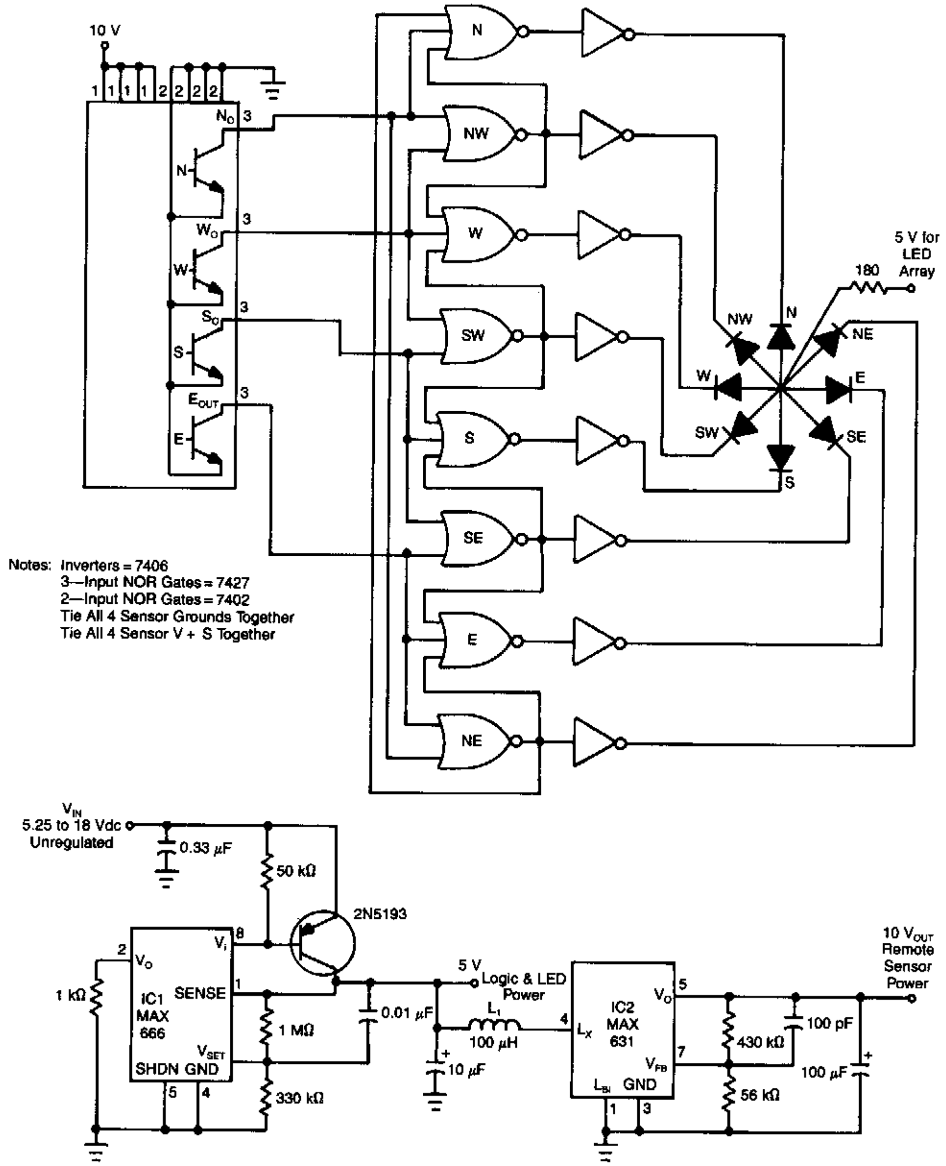
The output of this circuit will be a voltage that is equal to the peak of the input. D1 and C1 detect the peak voltage and this is read by the IC2 voltage follower.

25

Direction Finders

The sources of the following circuits are contained in the Sources section, which begins on page 665. The figure number in the box of each circuit correlates to the entry in the Sources section.

DIGITAL COMPASS



EDN

Fig. 25-1

A four output Hall sensor combined with a few logic gates produce this digital compass. The NOR gates resolve the four Hall outputs into eight distinct compass directions. LEDs to indicate direction are driven by eight inverters. A power supply for 5.25- to 18-Vdc operation is shown in the figure.

RADIO DIRECTION FINDER

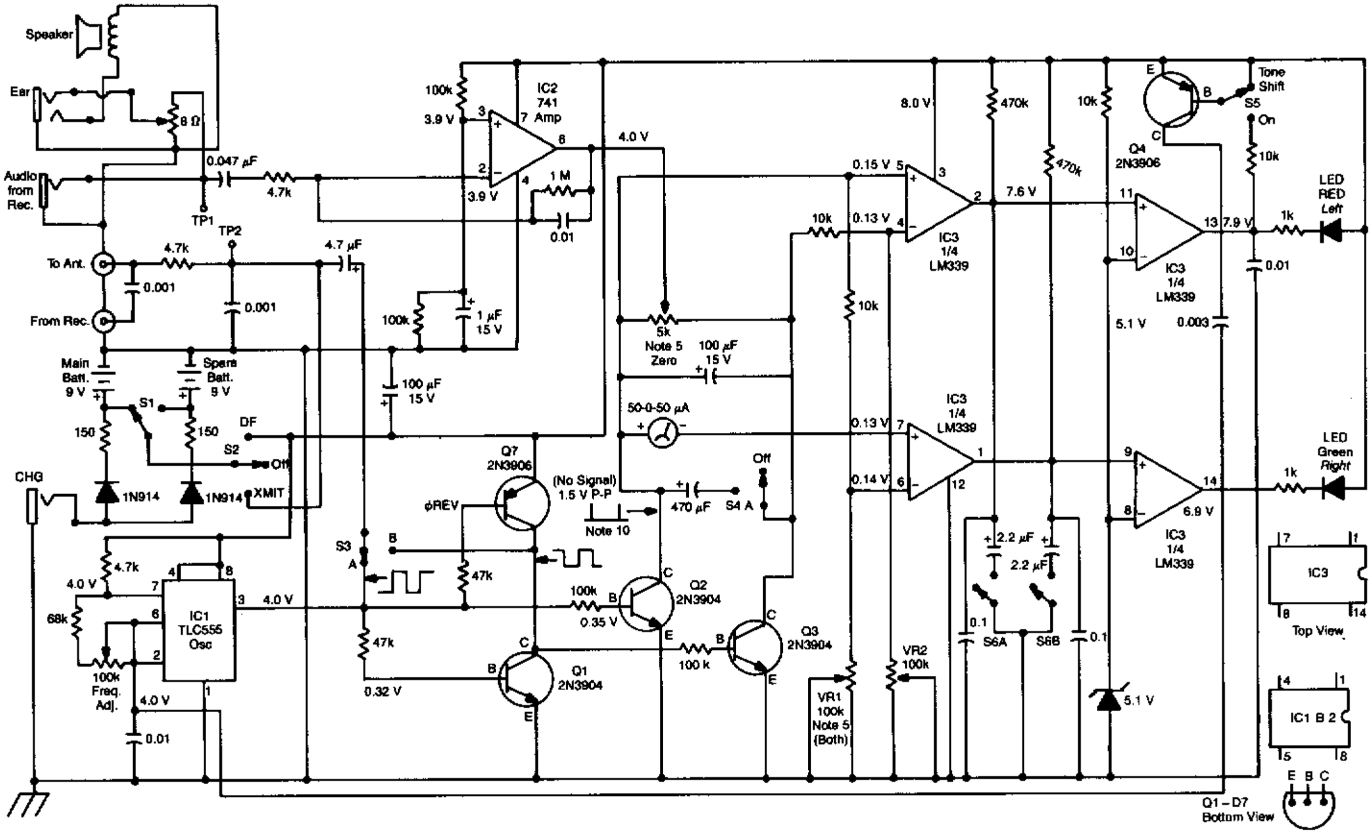
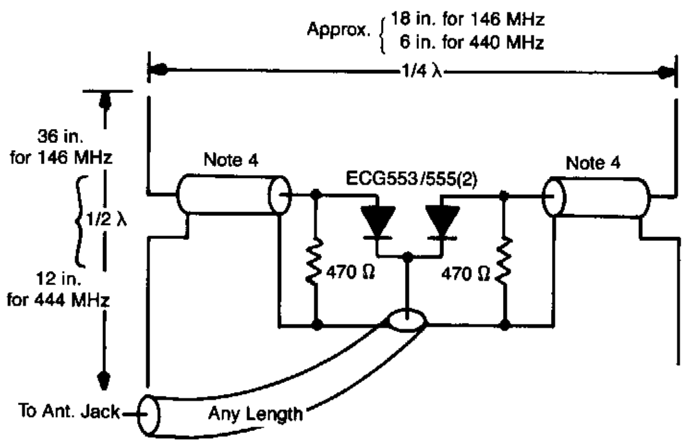
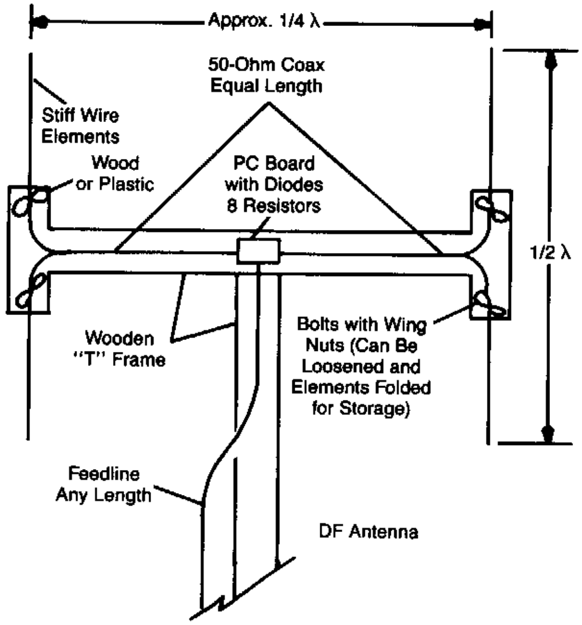


Fig. 25-2



Antenna construction.



Mechanical mounting details.

73 AMATEUR RADIO **Fig. 25-2**

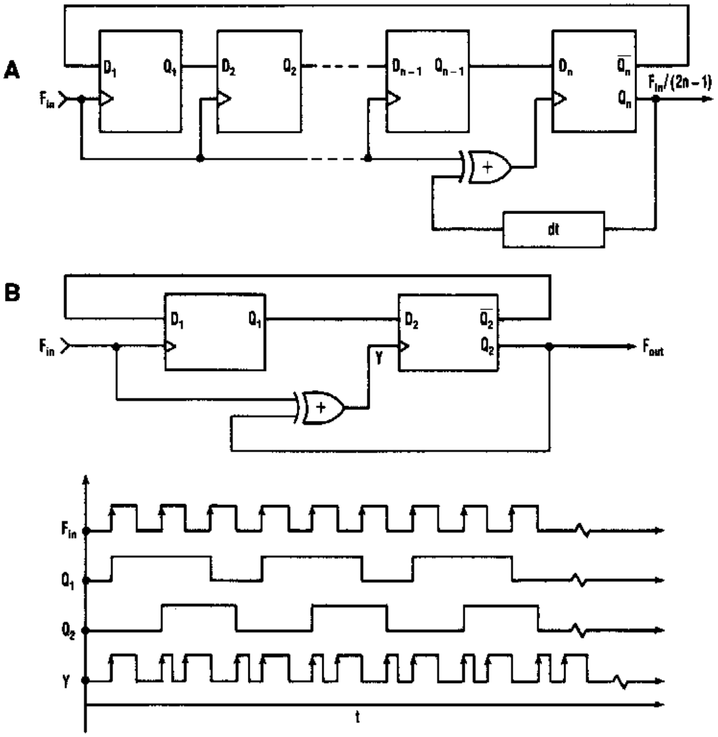
This RDF circuit consists of a square-wave oscillator (IC1), which switches two antennas alternately at an audio rate. A phase detector (Q1, 2, 3, 7) is used to compare receiver output amplified by IC2 with the reference phase from IC2 with the reference phase from IC1. A 50-μA meter is used as a left-right indicator. IC3 is a comparator used to drive indicator LEDs.

26

Dividers

The sources of the following circuits are contained in the Sources section, which begins on page 665. The figure number in the box of each circuit correlates to the entry in the Sources section.

CLOCK INPUT FREQUENCY DIVIDER



1. THE INPUT CLOCK frequency fed into this circuit is divided by 2n−1. The circuit consists of n clocked flip-flops and one exclusive-OR gate. The dt delay is zero in most cases.

2. THIS CIRCUIT CONFIGURATION divides the input frequency by three (a). The circuit's timing diagram verifies the division (b).

ELECTRONIC DESIGN

Fig. 26-1

ICA, R1 through R3, and Q1 form a current source. The current that charges C1 is given by:

$$I = \frac{(V_D \times R_1)}{(R_1 + R_2) \times R_3}$$

$$= \frac{(15 \times 3\ \text{k}\Omega)}{(3\ \text{k}\Omega + 12\ \text{k}\Omega) \times 470\ \text{k}\Omega}$$

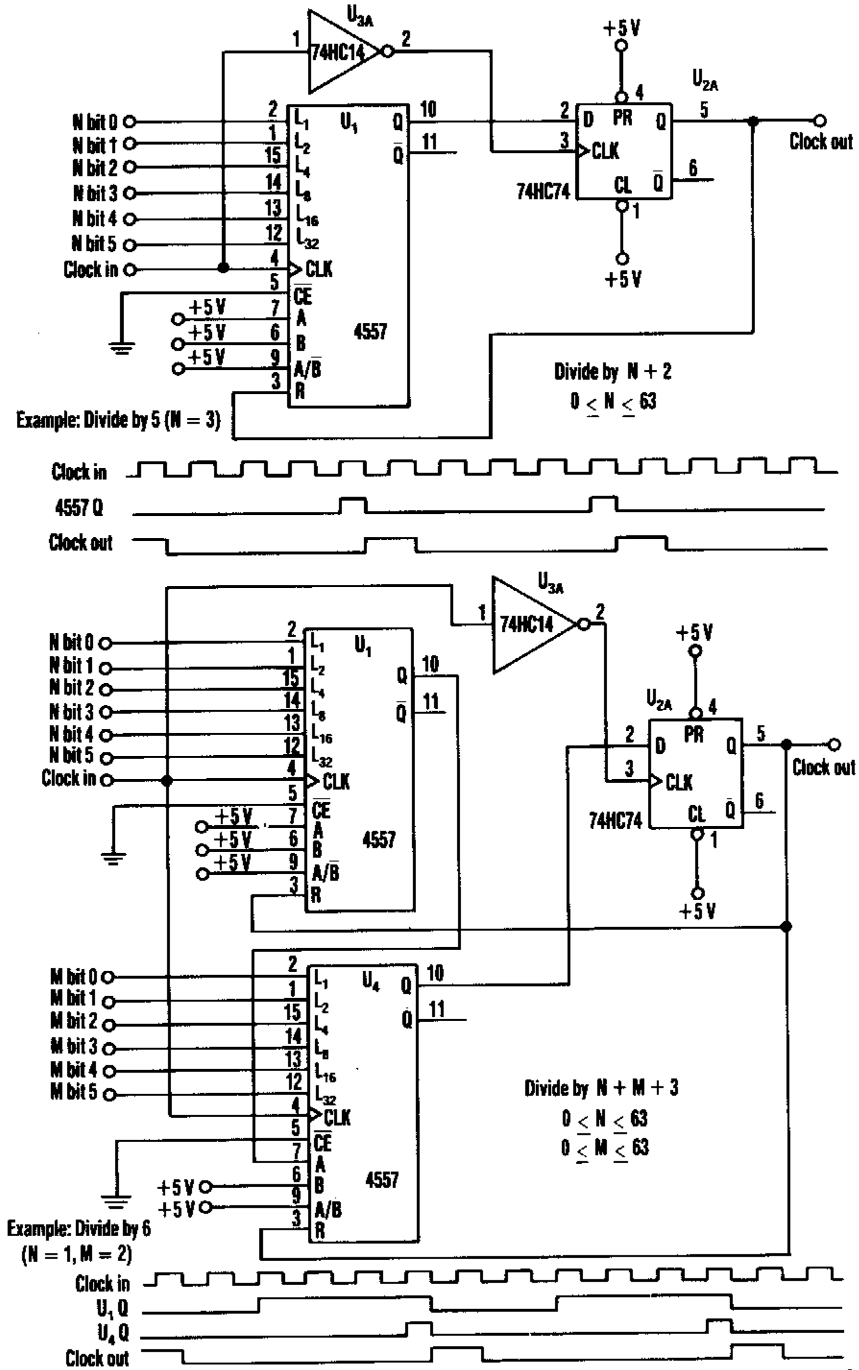
$$= 6.4\ \mu\text{A}$$

The input signal drives ICD. Because ICD's positive input (V+) is slightly offset to +0.1 V, its steady-state output will be near +13 V. This voltage is sent to ICC through D2, setting ICC's output to +13 V. Therefore, point D is cut off by D1, and C1 is charged by the current source. Assuming the initial voltage on C1 is zero, the maximum voltage (VC_{max}) is given by:

$$tw_{clk} > tp_{ff} + dt + tp_{xr} + tw_{ff}$$

The right side of the inequality should be the minimum pulse width (either up time or down time) of the input clock. The circuit, when constructed with standard 74F-type parts, operates without any added delay in the exclusive-OR feedback path and with an input frequency of up to 22.5 MHz. The circuit's output signal will have the same duty cycle as the input clock.

PROGRAMMABLE FREQUENCY DIVIDER



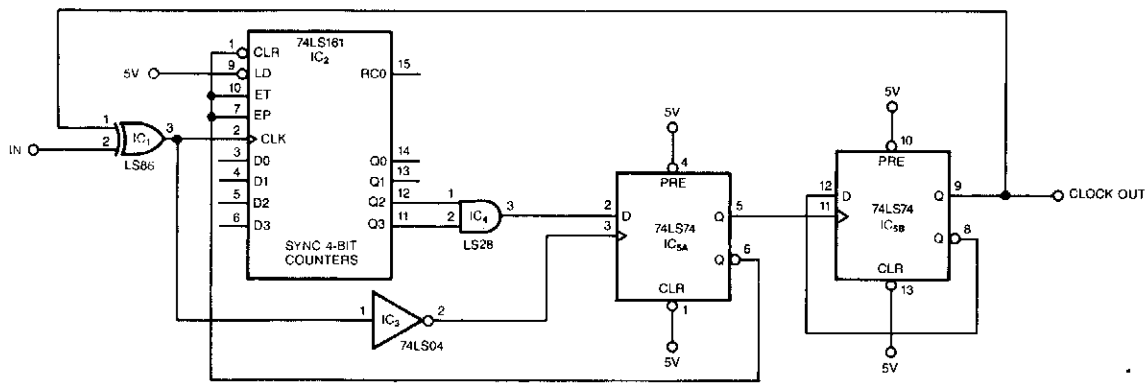
ELECTRONIC DESIGN

Fig. 26-2

PROGRAMMABLE FREQUENCY DIVIDER (*Cont.*)

This divider uses a variable-length shift register, a type-D flip-flop, and an inverter. The clock feeds the flip-flop clock input and the output of the shift register feeds the D input of the flip-flop. The FF output is tied back to the reset input of the shift register so that each clock pulse shifts a "1" into the 4557. $N+1$ cycles after the reset pulse is removed. The first "1" will propagate through the register output. The "1" is latched into the FF on the clock's next falling edge and fed back to the 4557 reset pin, which resets the shift register to zero. When a zero is clocked into the flip-flop on the next falling clock edge, the reset is removed, restarting the process. The divide ratio is $(N+2)$, where N = the binary number that is programmed into 4557.

DIVIDE-BY-ODD-NUMBER COUNTER



EDN

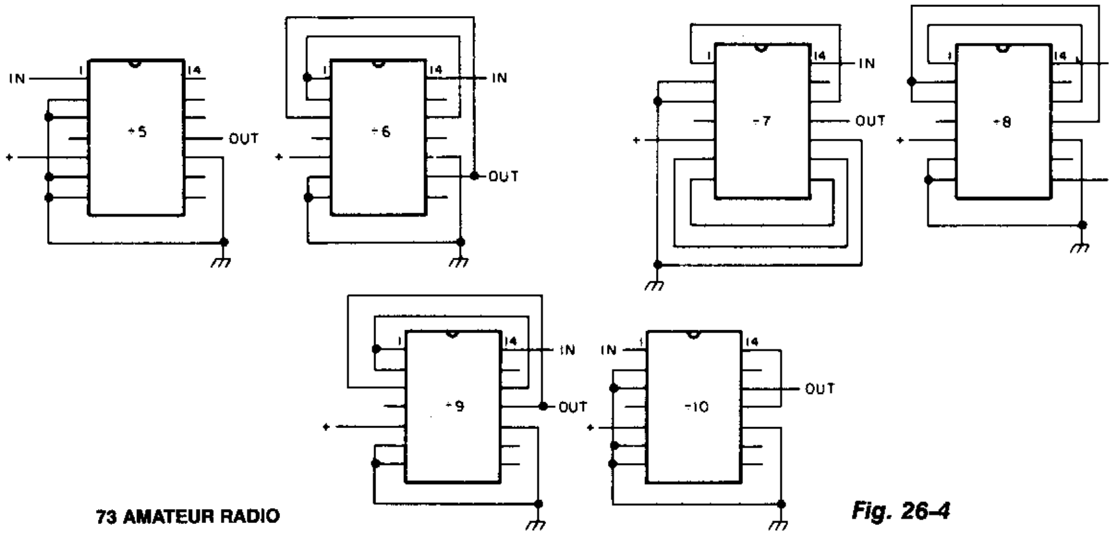
Fig. 26-3

This circuit symmetrically divides an input by virtually any odd number. The circuit contains $n+1/2$ clocks twice to achieve the desired divisor. By selecting the proper n, which is the decoded output of the 74LS161 counter, you can obtain divisors from 3 to 31. This circuit divides by 25; you can obtain higher divisors by cascading additional LS161 counters.

The counter and IC5A form the $n+1/2$ counter. Once the counter reaches the decoded counts, n, IC5A ticks off an additional 1/2 clock, which clears the counter and puts it in hold. Additionally, IC5A clocks IC5B, which changes the clock phasing through the XOR gate, IC1. The next edge of the input clocks IC5A, which reenables the counter to start counting for an additional $n+1/2$ cycles.

Although the circuit has been tested at 16 MHz, a worst-case timing analysis reveals that the maximum input frequency is between 7 and 8 MHz.

7490 ÷ *N* CIRCUITS

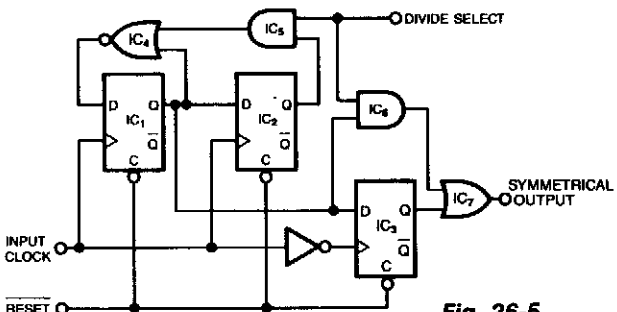


73 AMATEUR RADIO

Fig. 26-4

A 7490, 74LS90, 74C90, etc., is a decode divider, but it can be configured to divide by any *N* up to 10. The above figures illustrate the connections necessary to divide by *N* from 5 to 10.

DIVIDE-BY-2-OR-3 CIRCUIT



EDN

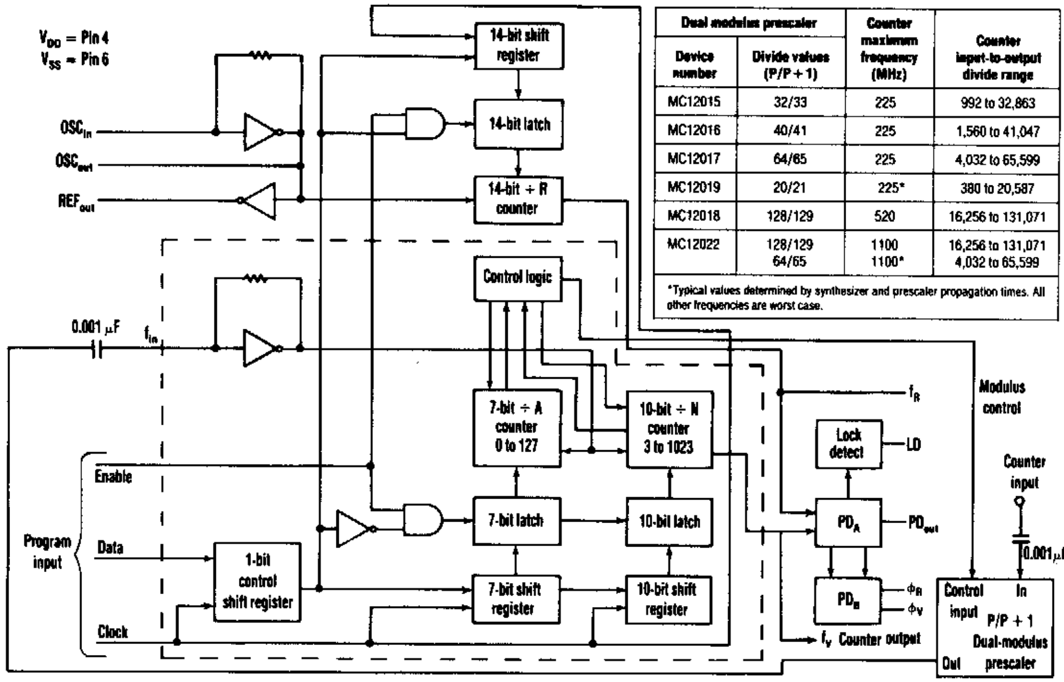
Fig. 26-5

This circuit produces a symmetrical waveform when dividing by either 2 or 3. The Divide Select input controls the division factor. When Divide Select is high, flip-flops IC1 and IC2, along with associated gates, form the classical divide-by-3 circuit.

When divide select is low, however, the output of the AND gate, IC5, goes low. Consequently, the NOR gate, IC4, inverts the feedback signal and passes it to the D input of the flip-flop, IC1. Now, IC1 acts like a toggle flip-flop and produces a divide-by-2 output.

IC3, which is, in effect, a negative-edge-triggered flip-flop, provides symmetrical output signals. When you select division by 2 (Divide Select is low), the output and AND gate IC6 is low, and IC3 simply clocks out the divider's output, delayed by one clock period. When you set Divide Select high, the path to the output through the AND and OR gates, IC6 and IC7, is enabled. This path means that the output goes high on the leading edge of IC3's input (not its output) and produces a symmetrical divide-by-3 output.

1 + -GHz DIVIDE-BY-*N* COUNTER



Dual modulus prescaler		Counter maximum frequency (MHz)	Counter input-to-output divide range
Device number	Divide values (P/P + 1)		
MC12015	32/33	225	992 to 32,863
MC12016	40/41	225	1,560 to 41,047
MC12017	64/65	225	4,032 to 65,599
MC12019	20/21	225*	380 to 20,587
MC12018	128/129	520	16,256 to 131,071
MC12022	128/129 64/65	1100 1100*	16,256 to 131,071 4,032 to 65,599

*Typical values determined by synthesizer and prescaler propagation times. All other frequencies are worst case.

ELECTRONIC DESIGN ***Fig. 26-6***

Counter speeds for CMOS- and TTL-programmable counters are limited to under 100 MHz. ECL-type devices can approach a few hundred MHz, but with significant current requirements. However, coupling the dual-modulus-prescaling technique with the available phase-locked-loop synthesizer chips that control the prescaler circumvents these frequency and power-drain constraints.

With this approach, designers can also choose various counter-programming schemes (serial, parallel, or data bus), in addition to achieving higher frequency capabilities. Low-power drain (less than 75 mW) and low-cost devices can also be selected. Moreover, only two ICs are necessary to achieve divide values above 131 000.

Maximum input frequency and dividing range for the counter are controlled by choosing an appropriate 8-pin dual-modulus prescaler. The counter's output appears at synthesizer pin F_V (see the figure). The total input-to-output divide value is governed by the equation:

$$N_{TOTAL} = N \times P + A$$

N and A represent the value programmed through the serial port into the divide-by-N and divide-by-A counters. P is the lower dual-modulus value that is established by the synthesizer's modulus-control signal.

Typically, A varies from zero to $P-1$ to achieve steps within the system's divide range. N must be equal to or greater than A. $N > A$ then sets a lower limit on N_{TOTAL}, which is dictated by $A_{MAX} = P-1$.

DIVIDE-BY-$N+1/2$ CIRCUIT

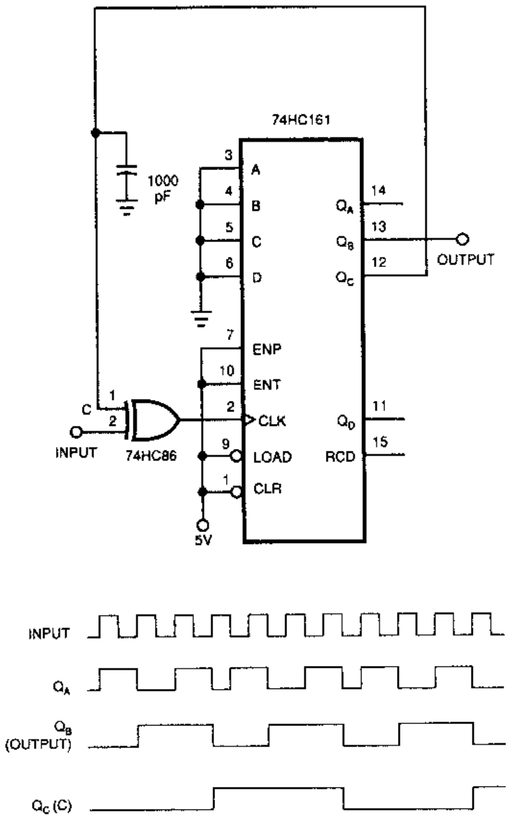


Table 1—XOR feedback signals for N+½ divider

Divide number	Feedback signal(s)			
N=1.5	Q_1			
N=2.5	Q_0	Q_2		
N=3.5	Q_2			
N=4.5	Q_0	Q_3		
N=5.5	Q_0	Q_1	Q_3	
N=6.5	Q_1	Q_3		
N=7.5	Q_3			
N=8.5	Q_0	Q_4		
N=9.5	Q_0	Q_2	Q_4	
N=10.5	Q_0	Q_1	Q_2	Q_4
N=11.5	Q_0	Q_1	Q_4	
N=12.5	Q_1	Q_4		
N=13.5	Q_1	Q_2	Q_4	
N=14.5	Q_2	Q_4		
N=15.5	Q_4			
N=16.5	Q_0	Q_5		
N=17.5	Q_0	Q_3	Q_5	
N=18.5	Q_0	Q_2	Q_3	Q_5
N=19.5	Q_0	Q_2	Q_5	
N=20.5	Q_0	Q_1	Q_2	Q_5

EDN *Fig. 26-7*

This circuit, instead of dividing by an integer, divides the input signal by $N+1/2$. With the feedback connections exactly as the figure shows, the circuit divides by 3.5. Point C ultimately controls when the input clocks the 74HC161 4-bit counter. When $C=0$, the positive edge of the input triggers the counter. If $C=1$, the negative edge of the input triggers the counter. Each time that point C changes level, the circuit shortens the output pulse width of the counter by half of an input cycle. Thus, the counter's divisor depends on how many changes occur at point C during one output period.

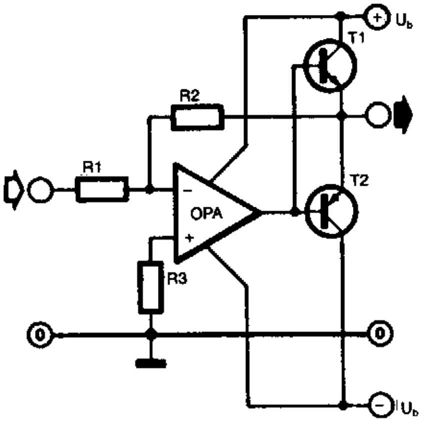
Although the figure divides by 3.5, feeding back different counter outputs produces different divisors. Generally, an m-bit binary counter with pure exclusive-OR (XOR) feedback can form an $N+1/2$ counter, where N ranges from $2^{m-2}+1/2$ to $2^{m-1}-1/2$. The divided output is available at the m-1 bit of the counter.

27

Driver Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 665. The figure number in the box of each circuit correlates to the entry in the Sources section.

OP AMP POWER DRIVER



ELEKTOR ELECTRONICS

Fig. 27-1(a)

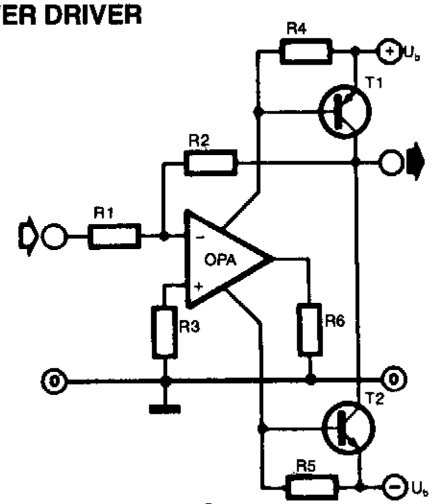


Fig. 27-1(b)

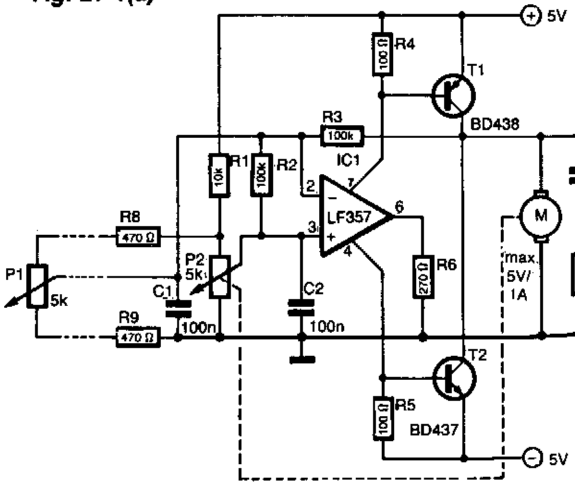


Fig. 27-1(c)

Frequently, the output current of an operational amplifier is inadequate for the application as, for instance, when a small motor or loudspeaker has to be driven. Normally, this is resolved by adding an emitter follower to the circuit as shown in Fig. 27-1(a). Unfortunately, that circuit does not allow the full supply voltage, U_b, to be used, because the output voltage of the op amp must always be 1 to 2 V smaller than $\pm U_b$. To that must be added the drop across the base-emitter junction of transistors T1 and T2.

The circuit shown in Fig. 27-1(b) (principle) and Fig. 27-1(c) (practical) is a more appropriate solution: it was designed specifically for driving small motors. Since the output current of the op amp flows through its supply lines, the driver transistors may also be controlled over these lines.

The value of base-emitter resistors R4 and R5 has been chosen to ensure that in spite of the quiescent current through the op amp, T1 and T2 are switched off. Resistor R6 limits the output current of the op amp. If the op amp is a type with guaranteed short-circuit protection, R6 may be replaced by a jump lead.

The output voltage is only 50 to 100 mV (collector-emitter saturation voltage of the driver transistors) smaller than the supply voltage. When choosing these transistors, it is therefore essential to take into account the saturation voltage (in addition to the maximum current amplification and power rating).

The value of the resistors in an inverting circuit is calculated from:

$$\alpha = \frac{R_2}{R_1}$$

and:

$$R_3 \approx \frac{R_2}{R_1}$$

where α is the amplification.

In a noninverting circuit (R1 between the −input and ground and the input signal connected to the +input of the op amp), the amplification is:

$$\alpha = \frac{R_2}{(R1+1)}$$

and:

$$R_3 < < R_e$$

$$R_4 < \frac{+\alpha}{=U_b}$$

$$R_5 < \frac{-0.5\alpha}{U_b}$$

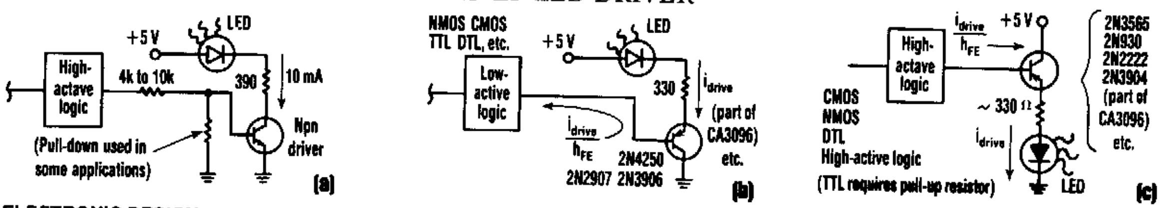
$$R_6 \approx \frac{U_b}{I_{max}}$$

where R_e is the input impedance of the op amps.

The circuit can be used with discrete (single) op amps only, because double or quadruple types in one package share the supply voltage pins. The setting accuracy of the circuit in Fig. 27-1(c) is better than 1%.

EMITTER/FOLLOWER LED DRIVER

TYPICAL NPN LED DRIVER   PNP EF LED DRIVER   NPN EF DRIVER

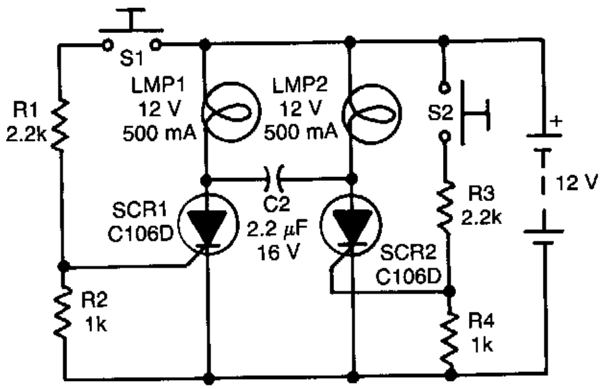


ELECTRONIC DESIGN

Fig. 27-2

Using emitter/followers saves parts and simplifies LED driver circuits and generally produces less loading on logic circuitry.

FLIP-FLOP INDEPENDENT LAMP DRIVER



RADIO-ELECTRONICS

Fig. 27-3

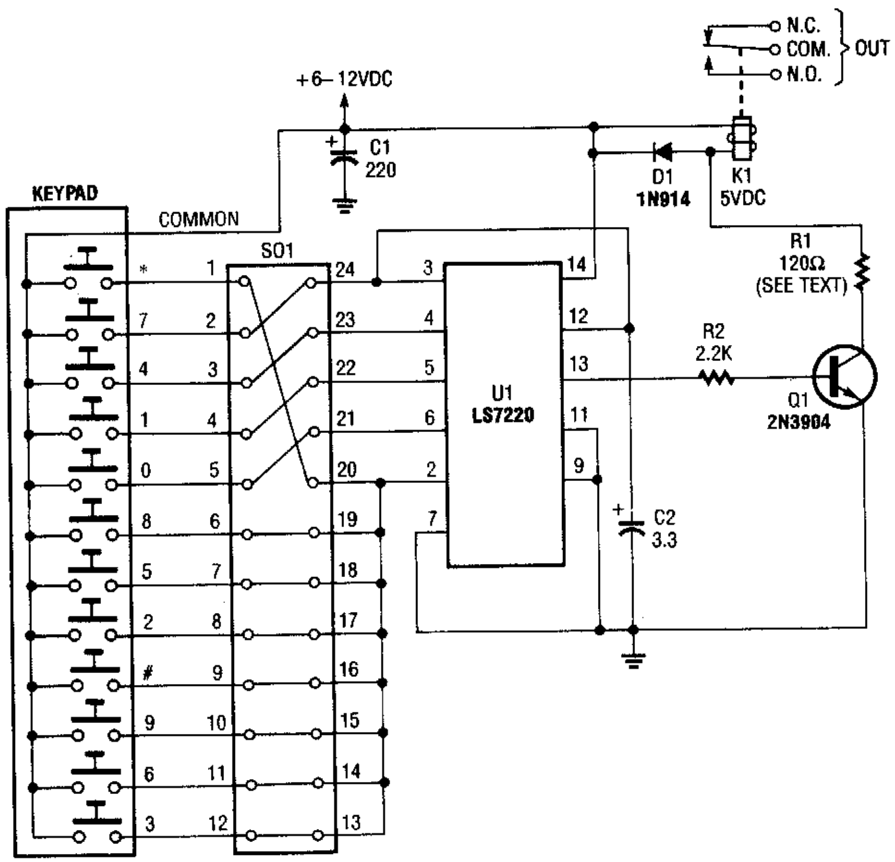
Assume first that SCR1 is on and SCR2 is off so that C1 is fully charged, with its LMP2 end positive. The state of the circuit can be changed by pressing S2. As SCR2 turns on, it turns SCR1 off capacitively via its anode. Capacitor C1 then recharges in the opposite manner (i.e., the left end is now positive). The state of the circuit can be changed again by pressing S1, thus driving SCR1 on by way of its gate, and driving SCR2 off capacitively via its anode.

28

Electronic Locks

The sources of the following circuits are contained in the Sources section, which begins on page 665. The figure number in the box of each circuit correlates to the entry in the Sources section.

DIGITAL ENTRY LOCK

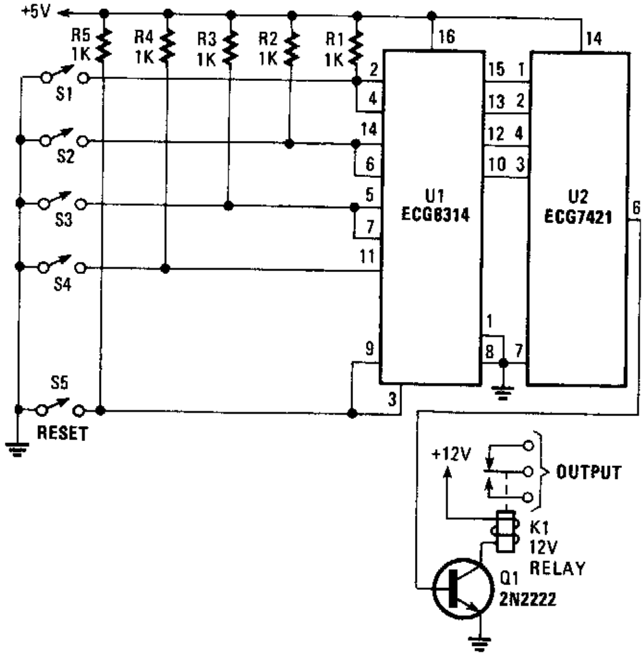


POPULAR ELECTRONICS

Fig. 28-1

A keypad enters a four-digit access code, which is programmed via jumpers on a 24-pin plug-in header and socket. U1 is an LST220, which detects a four-digit sequential data input. When the correct data is entered into the keyboard, pin 13 of U1 goes high, which activates Q1 and K1. K1 drives an external electric lock solenoid, etc.

KEYLESS LOCK



POPULAR ELECTRONICS *Fig. 28-2*

The circuit uses a four-bit latch (U1). What makes the circuit sequential is that the set input of the first bit latch is tied to the reset of the second bit latch, and so forth. That ensures that any bit latched will be reset by the previous bit latch. The ECG8314 also has a master reset (pin 9) that is tied to the first bit-latch reset (pin 3), which provides an added measure of security for the lock.

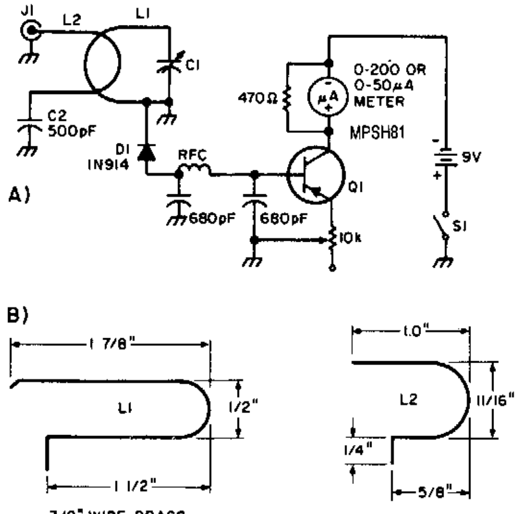
The outputs of U1 are fed to a four-input AND gate (U2), then to Q1 (used as switching transistor), which is used to drive relay K1. The EGC8314 has an enable low (pin) that can be used as a timing circuit, if that is desired.

29

Field-Strength Meters

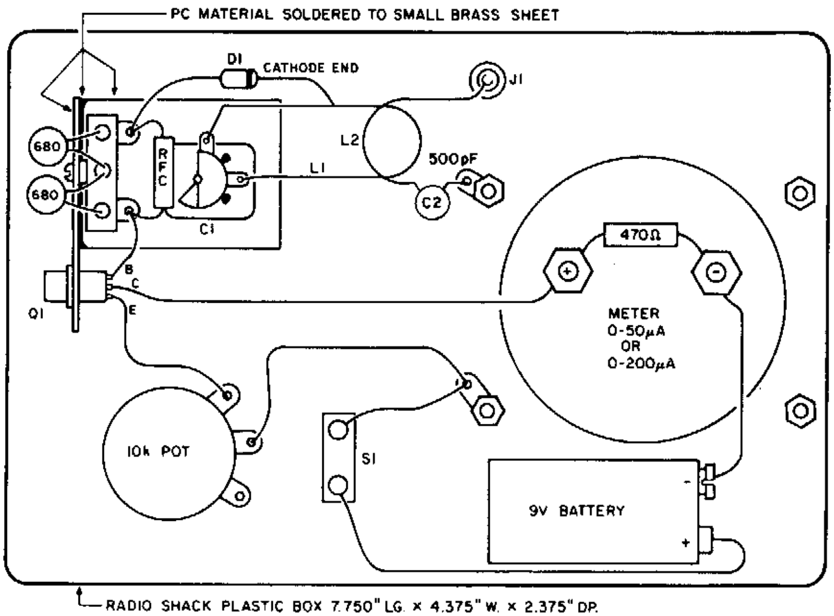
The sources of the following circuits are contained in the Sources section, which begins on page 665. The figure number in the box of each circuit correlates to the entry in the Sources section.

UHF FIELD-STRENGTH METER



AMATEUR RADIO **_Fig. 29-1(a)_**

Field strength meter schematic.

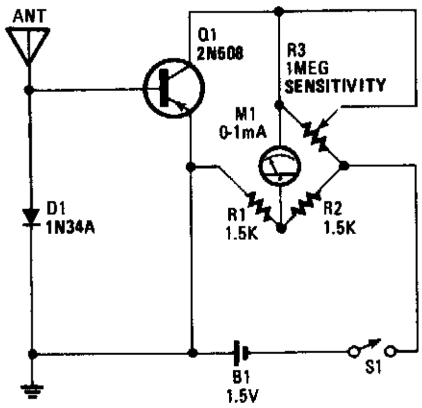


AMATEUR RADIO **_Fig. 29-1(b)_**

Parts layout for the field strength meter.

Useful for transmitter or antenna alignment, this meter covers 400 to 500 MHz. An amplifier stage is included for improved sensitivity. Follow the layout in Fig. 29-1(b).

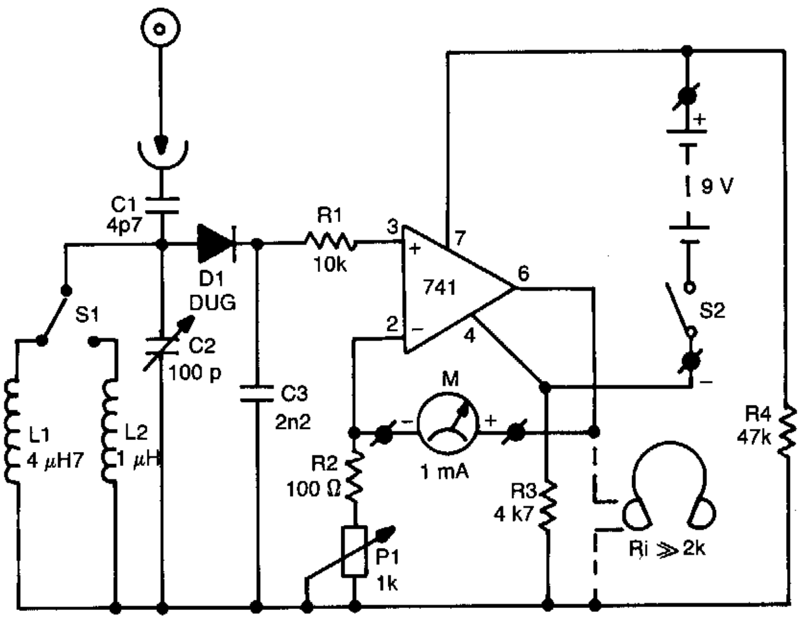
FIELD-STRENGTH METER



POPULAR ELECTRONICS *Fig. 29-2*

This field-strength meter is basically a bridge circuit that is equipped with a 0-to-1-mA meter as a readout.

SIGNAL-STRENGTH METER



ELEKTOR ELECTRONICS *Fig. 29-3*

This field-strength meter is useful for antenna testing. It covers 6 to 60 MHz and uses a rugged 0-to-1 mA meter. A 9-V battery supplies power. The unit can be mounted in a small plastic or metal case.

30

Filter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 665. The figure number in the box of each circuit correlates to the entry in the Sources section.

Fast-Response (settling) Low-Pass Filter
Tunable Audio Filter
Turntable Rumble Filter
Tunable Bandpass Filter
Low-Cost Crystal Filters
Antialiasing and Sync-Compensation Filter
Two-Section 300-to-3 000 Hz Speech Filter
300-to-3 400 Hz Second-Order Speech Filter
Fourth-Order 100-Hz High-Pass Filter
Simple Ripple Suppressor
Second-Order 100-Hz High-Pass Filter
Scratch Filter
Simple Rumble Filter
1 000:1 Tuning Voltage-Controlled Filter
Two Sallen-Key Low-Pass Active Filter

FAST-RESPONSE (SETTLING) LOW-PASS FILTER

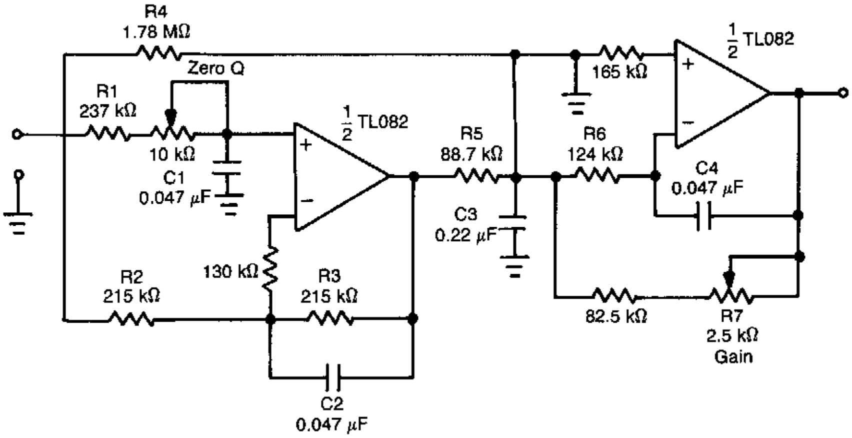


Fig. 30-1(a)

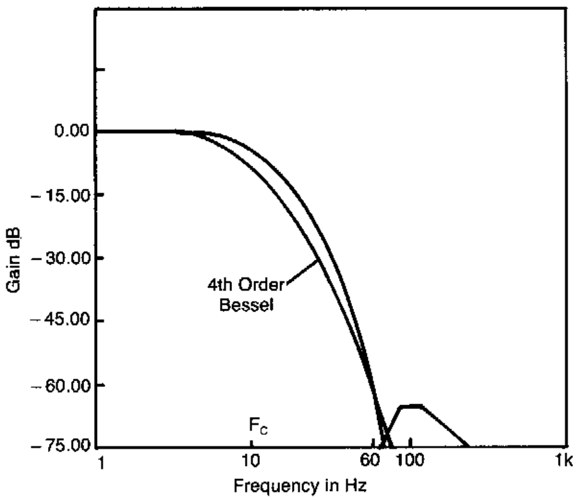


Fig. 30-1(b)

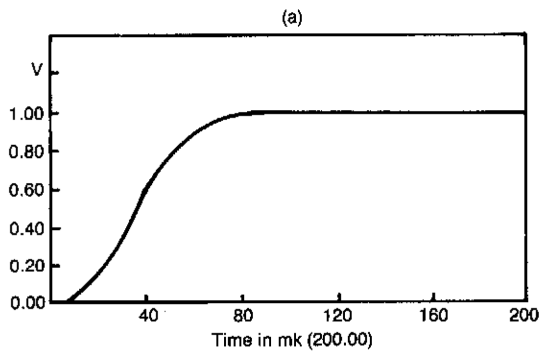


Fig. 30-1(c)

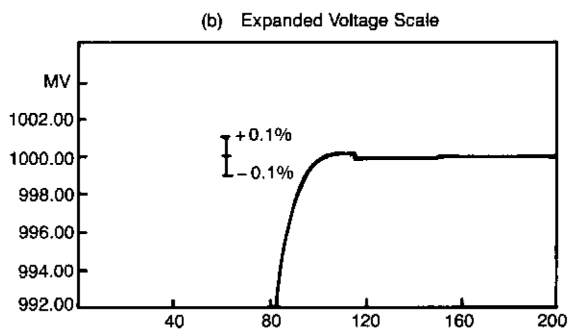


Fig. 30-1(d)

By introducing an extra transmission zero to the stopband of a low-pass filter, a sharp roll-off characteristic can be obtained. The filter design example of Fig. 30-1(a) shows that the time-domain performance of the low-pass section can also be improved. Figure 30-1(b) shows the attenuation characteristic of the proposed circuit. Position of the transmission zero is determined by the passive components around the first op amp. It was chosen to obtain 60 dB of rejection at 60 Hz.

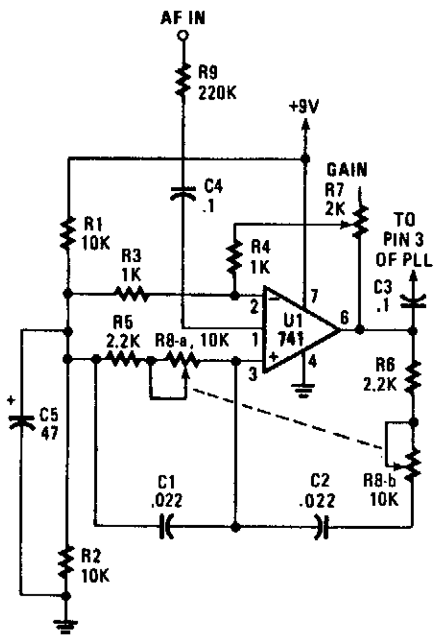
A suitable fourth-order Bessel filter has the frequency response, as shown by the dashed line. Its response to a step input is characterized by settling time to 0.1% of $1.8 \div F_C = 180$ ms.

Figure 30-1(c) and 30-1(d) represent the step response for the filter of Fig. 30-1(a) in both normal and expanded voltage scales. As you can see, settling time to 0.1% is below 100 ms; overshoot and ringing stay below 0.03%.

This quite significant speed and accuracy improvement can be a major factor, particularly for low-frequency applications. Averaging filter for low-frequency linear or true rms ac-to-dc converters is an example. Some anti-aliasing applications can also be considered.

For best results, resistance ratios $R_4 \div R_5 = 20$, $R_6 \div R_5 = 1.4$, and capacitance ratios $C_3 \div C_2 = C_3 \div C_4 = 4.7$ should be kept up for any selected F_C.

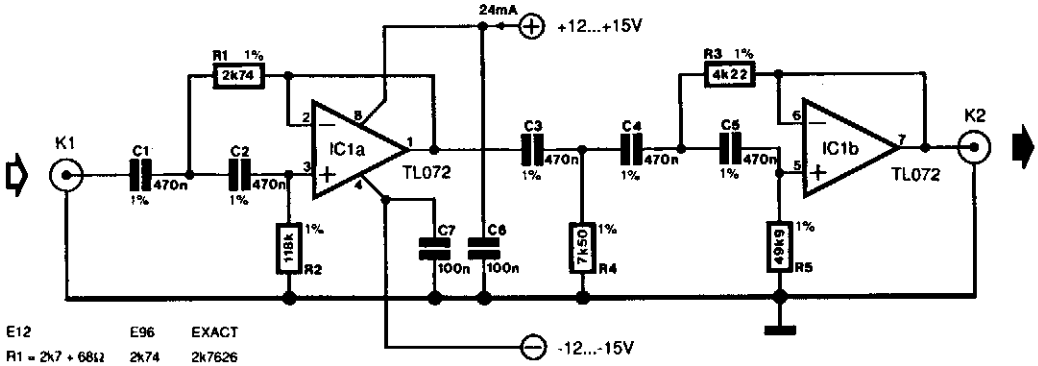
TUNABLE AUDIO FILTER



POPULAR ELECTRONICS *Fig. 30-2*

This circuit uses a Wien Bridge and variable negative feedback. R7 controls the gain and R8A and R8B controls the tuned frequency.

TURNTABLE RUMBLE FILTER



E12	E96	EXACT
R1 = 2k7 + 68Ω	2k74	2k7626
R2 = 120k	118k	119k05
R3 = 3k9 + 330Ω	4k22	4k2314
R4 = 6k8 + 680Ω	7k50	7k4648
R5 = 47k + 2k2	49k9	49k475

ELEKTOR ELECTRONICS *Fig. 30-3*

Many record players unfortunately exhibit two undesired side effects: rumble (noise caused by the motor and the turntable) and other low-frequency spurious signals. The active high-pass Chebyshev filter presented here was designed to suppress those noises. The filter has a 0.1-dB ripple characteristic and a cut-off point of 18 Hz.

The choice of a Chebyshev filter might not seem optimum for audio purposes, but because of its 0.1-dB ripple in the pass band it behaves very much like a Butterworth type. Its advantage is that the response has steeper skirts (which are calculated curves). Frequencies below 10 Hz are attenuated by more than 35 dB. The phase behavior in the pass band shows a gradual shift so that its effect on the reproduced sound is inaudible.

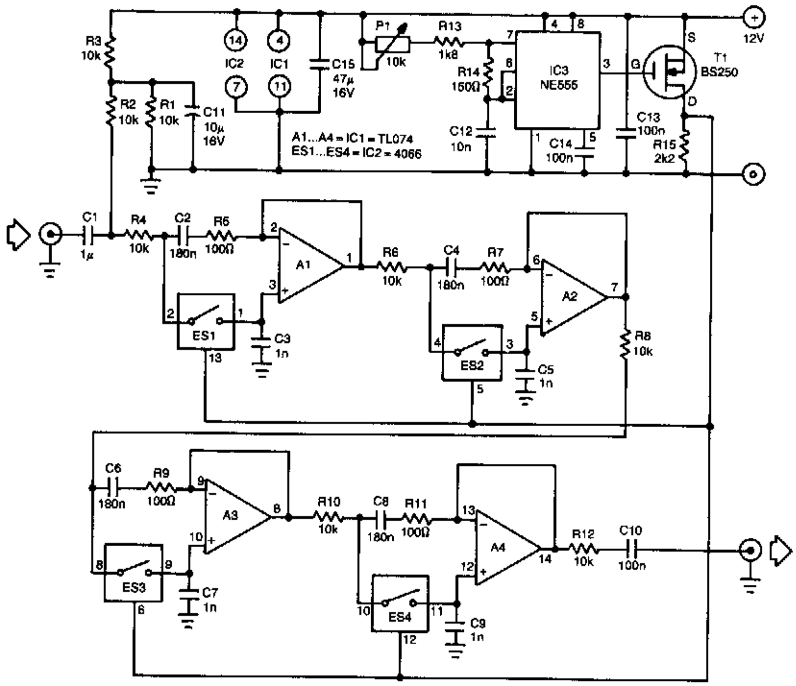
If the filter is used in a stereo installation, the characteristics of both filters must be identical or nearly so. Phase differences between channels can be heard—perhaps not so much at lower frequencies, but certainly in the mid ranges. To ensure identity and also to obtain the desired characteristics, capacitors C1 through C5 must be selected carefully. It does not matter much whether their value is 467 or 473 pF; this difference only causes a slight shift of the cut-off point. However, they must be identical within that 1% tolerance. For symmetry of channels, the capacitors can be paired and then used in either channel at the corresponding position.

The diagram shows theoretical values for the resistors: their practical values are given in the table. The prototype was constructed with 5% metal-film types from the E12 series and these were used without sorting. Their tolerance was perfectly acceptable in practice.

The current drawn by the circuit is purely that through the op amp and it amounts to about 4 mA. The high cut-off point is also determined by the op amp and it lies at about 3 MHz.

The only problem that cannot be foreseen is a possible coupling capacitor in the signal source. That component will be in series with C1 and this might adversely affect the frequency response. However, if its value is greater than 47 μF, it will have little if any effect; if it is below that value, it is best removed; C1 will assume its function.

TUNABLE BANDPASS FILTER



ELEKTOR ELECTRONICS

Fig. 30-4

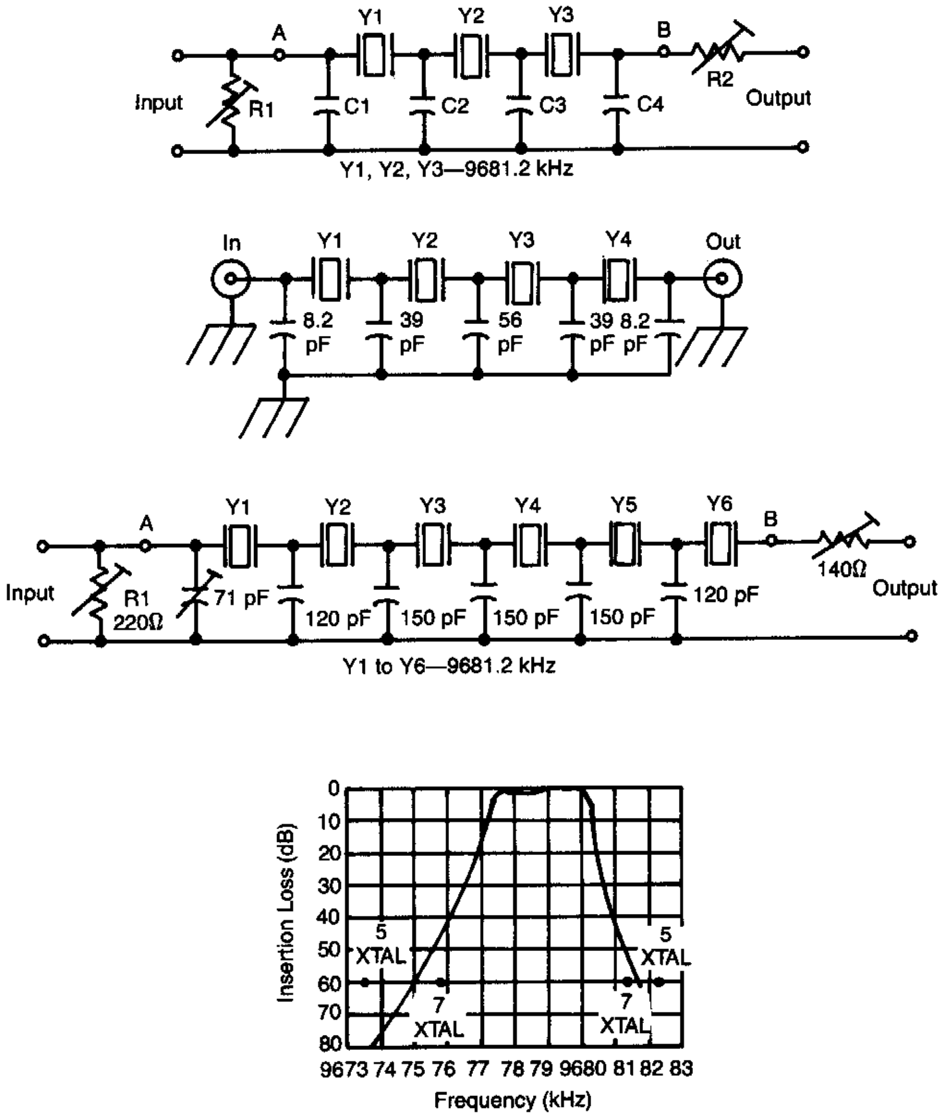
One of the difficulties in the design of higher-order tunable bandpass filters is achieving correct tracking of the variable resistors in the RC networks. The use of switched capacitor networks can obviate that difficulty, as is shown in this filter.

The filter can be divided roughly into two stages: an oscillator that controls the electronic switches and the four phase-shift networks that provide the filtering proper. The oscillator, based on a 555, generates a pulsating signal whose frequency is adjustable over a wide range: the duty factor varies from 1:10 to 100:1.

Electronic switches ES1 through ES4 form the variable resistors whose value is dependent on the frequency of the digital signal. The operation of these switches is fairly simple. When they are closed, their resistance is about 60 Ω; when they are open, it is virtually infinitely high. If a switch is closed for, say, 25% of the time, its average resistance is therefore 240 Ω. Varying the open:closed ratio of each switch varies the equivalent average resistance. The switching rate of the switches must be much greater than the highest audio frequency to prevent audible interference between the audio and the clock signals.

The input signal causes a given direct voltage across C1, so the op amp can be operated in a quasisymmetric manner, in spite of the single supply voltage. The direct voltage is removed from the output signal by capacitor C10.

The fourth-order filter in the diagram can be used over the entire audio range and it has an amplification of about 40, although this depends to some extent on the clock frequency. The bandwidth depends mainly on the set frequency. The circuit draws a current of not more than 15 mA.

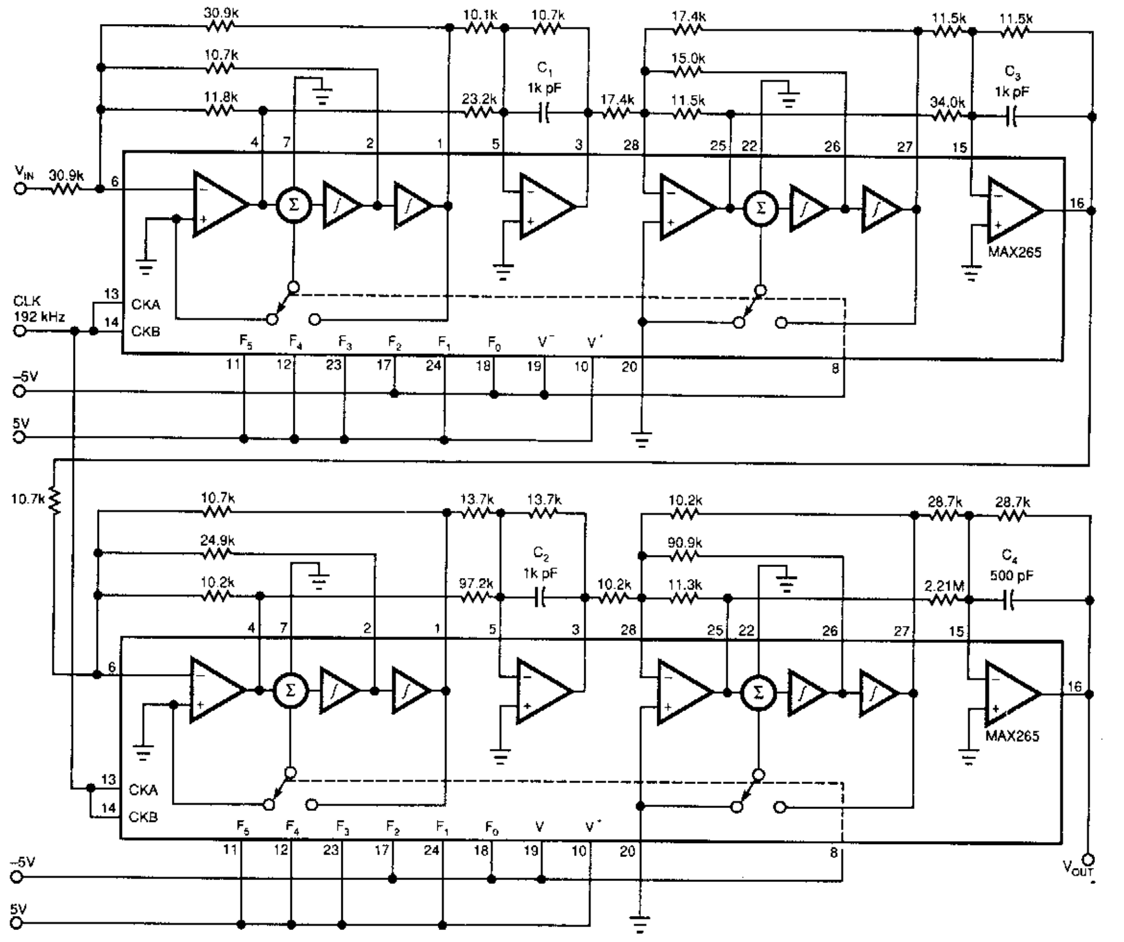


ARRL HANDBOOK

Fig. 30-5

Low-cost CB crystals can be used for these 9-MHz crystal ladder filters. Notice that the 27-MHz crystals (3rd overtone) are used on their fundamental frequencies.

ANTIALIASING AND SYNC-COMPENSATION FILTER



EDN

Fig. 30-6

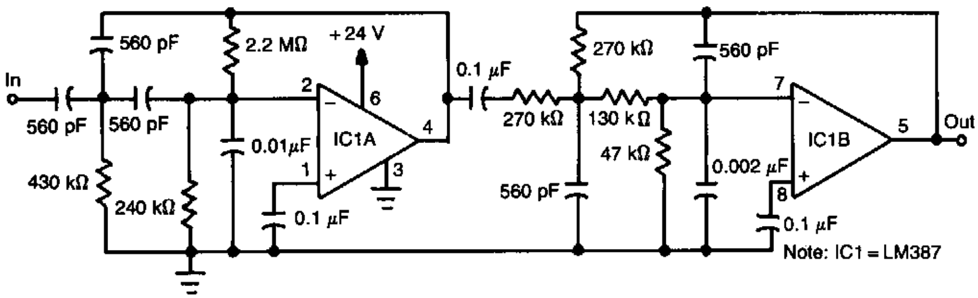
Two dual-biquad filter chips and some external components form a multipurpose filter to reconstruct D/A converter signals. Connected to a converter's output, the filter provides antialiasing, reduces the D/A converter's quantization noise, and compensates for $\sin(\pi x) \div (\pi x)$—the "sync" function (attenuation).

The circuit incorporates an inverse-sync function that operates to one-third of the converter's sample rate. Beyond one-third, the filter's response shifts to a stopband filter, which provides −70 dB attenuation. This attenuation conforms to the converter's inherent signal-to-noise ratio and quantization error.

To prevent aliasing, the stopband edge must be no higher than the Nyquist frequency ($f_{sn} \div 2$). To achieve 70-dB stopband rejection with this eighth-order filter requires a transition ratio ($f_{\text{STOPBAND}} \div f_{\text{PASSBAND}}$) of 1.5, which sets the passband's upper limit at $f_s \div 3$.

Notice also that you can apply a simple divide-by-64 circuit to the 192-kHz clock frequency to set the necessary 3× ratio between the converter's sample rate and the filter's 1-kHz corner frequency. The V^+, V^-, and the F_0 through F_5 connections program each filter chip for an f_{CLK}/f_0 ratio of 191.64.

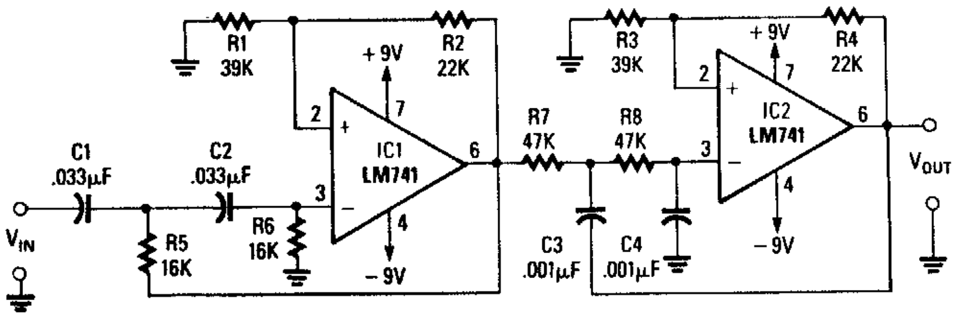
TWO-SECTION 300-3 000 Hz SPEECH FILTER



RADIO-ELECTRONICS **_Fig. 30-7_**

An LM387 dual low-noise amplifier is used in an active filter. Both sections are used to produce second-order HP and LP filters, respectively.

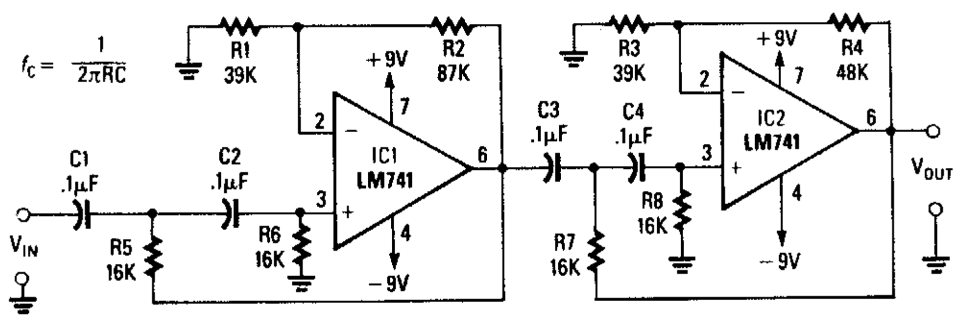
300-to-3 400 Hz SECOND-ORDER SPEECH FILTER



RADIO-ELECTRONICS **_Fig. 30-8_**

Using two op amps, this filter is designed for second-order response. It has a bandpass of 300 to 3 400 Hz, for applications in speech or telephone work.

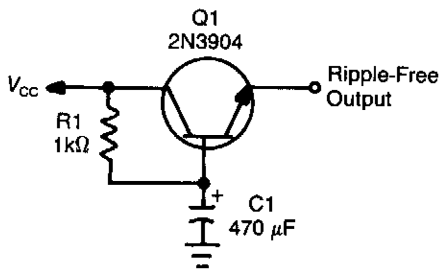
FOURTH-ORDER 100-Hz HIGH-PASS FILTER



RADIO-ELECTRONICS **_Fig. 30-9_**

This filter, using two sections of LM741, can be scaled in frequency, if desired.

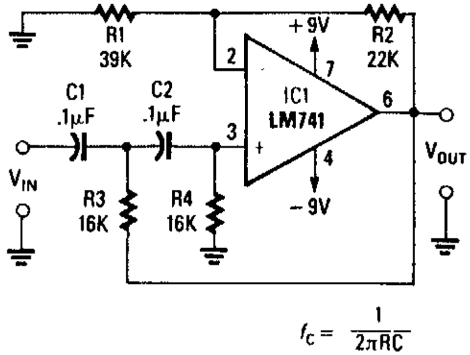
SIMPLE RIPPLE SUPPRESSOR



WELS' THINK TANK *Fig. 30-10*

This circuit, at times called a *capacitance multiplier*, is useful for suppression of power-supply ripple. C1 provides filtering that is equal to a capacitor of $(B+1)\,C_1$, where B = dc current gain of Q1 (typically > 50).

SECOND-ORDER 100-Hz HIGH-PASS FILTER

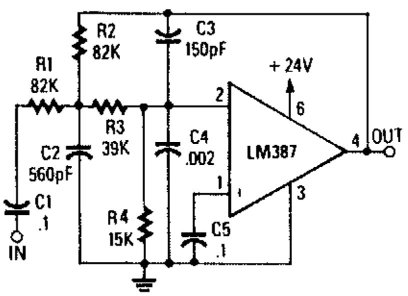


$$f_c = \frac{1}{2\pi RC}$$

RADIO-ELECTRONICS *Fig. 30-11*

This second-order filter can be scaled to change the cutoff frequencies.

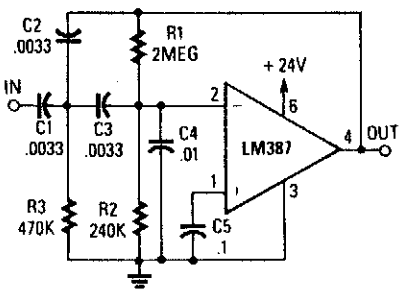
SCRATCH FILTER



RADIO-ELECTRONICS *Fig. 30-12*

Designed to produce 12-dB/octave roll-off above the 10-kHz cutoff frequency, this LP active filter will help reduce needle scratch on records. It uses an LM387 low-noise amplifier IC.

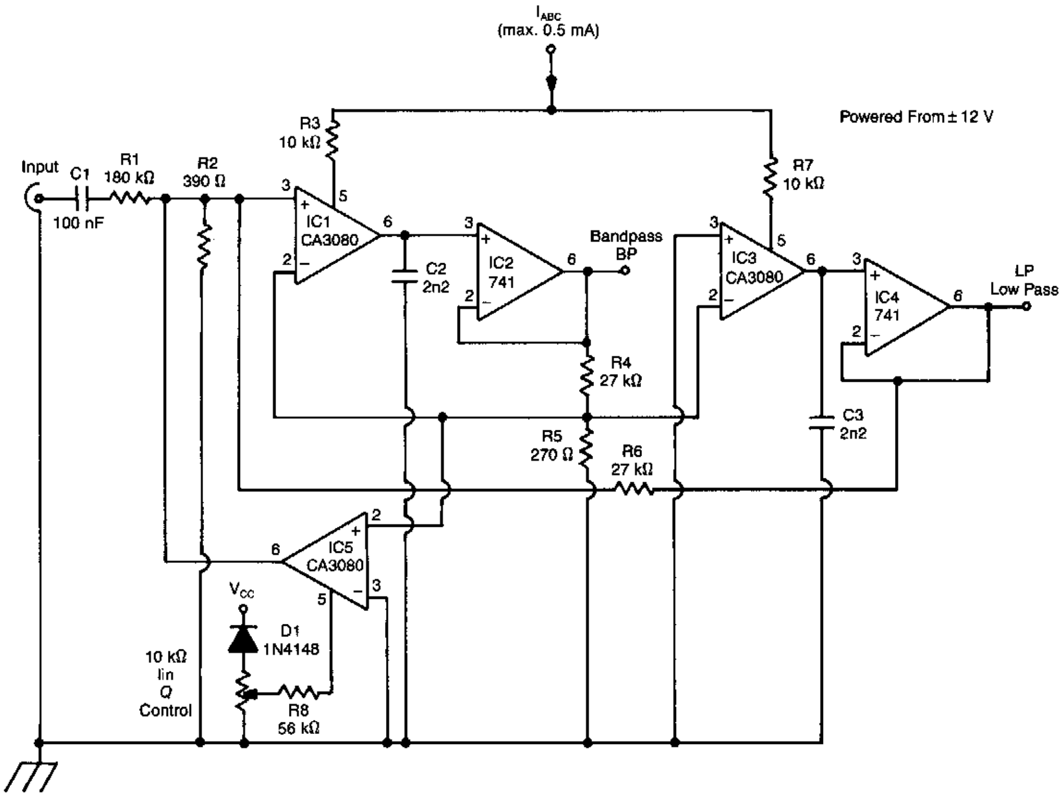
SIMPLE RUMBLE FILTER

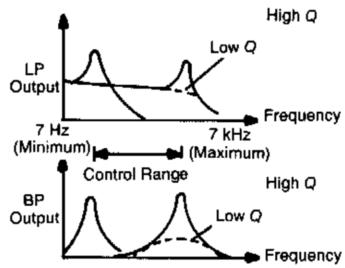


RADIO-ELECTRONICS *Fig. 30-13*

This circuit is a two-section active HP filter using an LM387, with a cutoff below 50 Hz at 12-dB per octave. It will help reduce rumble as a result of turntable defects in record systems.

1 000:1 TUNING VOLTAGE-CONTROLLED FILTER

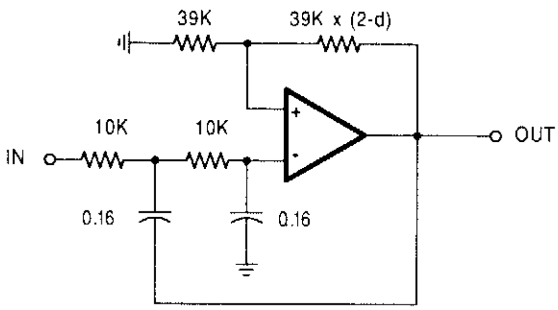




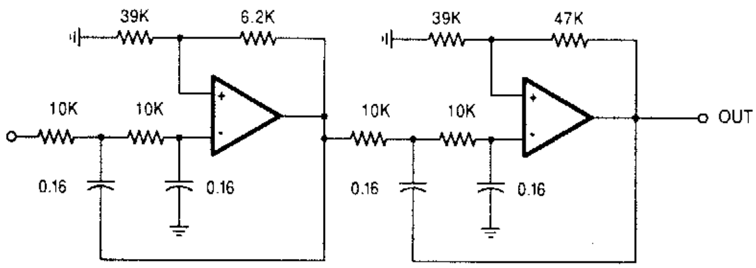
ELECTRONICS TODAY INTERNATIONAL ***Fig. 30-14***

A standard dual integrator filter can be constructed using a few CA3080s. By varying IABC, the resonant frequency can be swept over a 1 000:1 range. At IC1, three are current-controlled integrators. At IC2, four are voltage followers that serve to buffer the high-impedance outputs of the integrators. A third CA3080 (IC5) is used to control the Q factor of the filter. The resonant frequency of the filter is linearly proportional to I_{ABC}. Hence, this unit is very useful in producing electronic music. Two outputs are produced: a low-pass and a bandpass response.

TWO SALLEN-KEY LOW-PASS ACTIVE FILTERS



(A) SIMPLE SECOND ORDER SALLEN-KEY SECTION



(B) FOURTH ORDER BUTTERWORTH LOW PASS AUDIO FILTER

RADIO-ELECTRONICS ***Fig. 30-15***

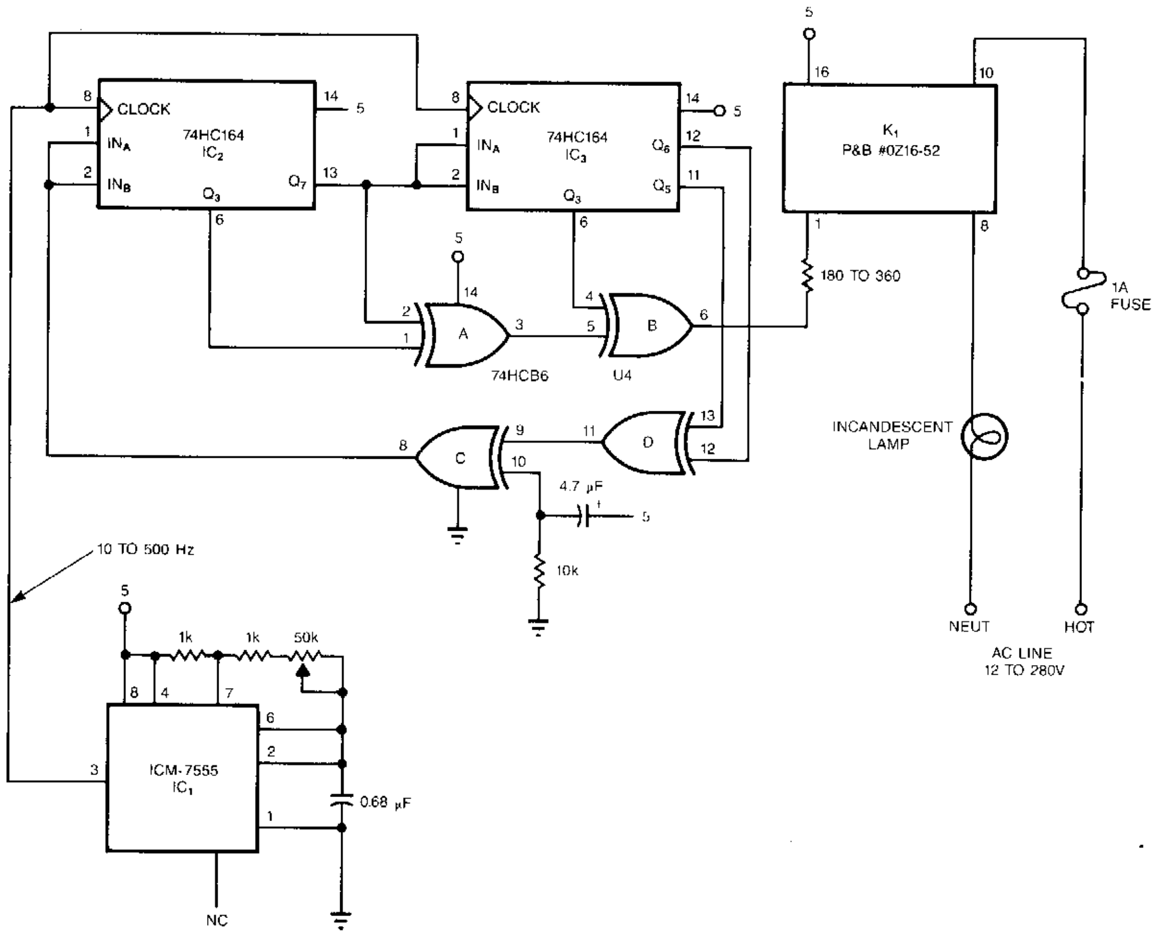
These filters are designed for 10-kΩ impedance level and 1-kHz cutoff frequency, but the components can be scaled as required for other impedances and cutoff frequencies.

31

Flashers and Blinkers

The sources of the following circuits are contained in the Sources section, which begins on page 666. The figure number in the box of each circuit correlates to the entry in the Sources section.

PSEUDORANDOM SIMULATED FLICKER SEQUENCER



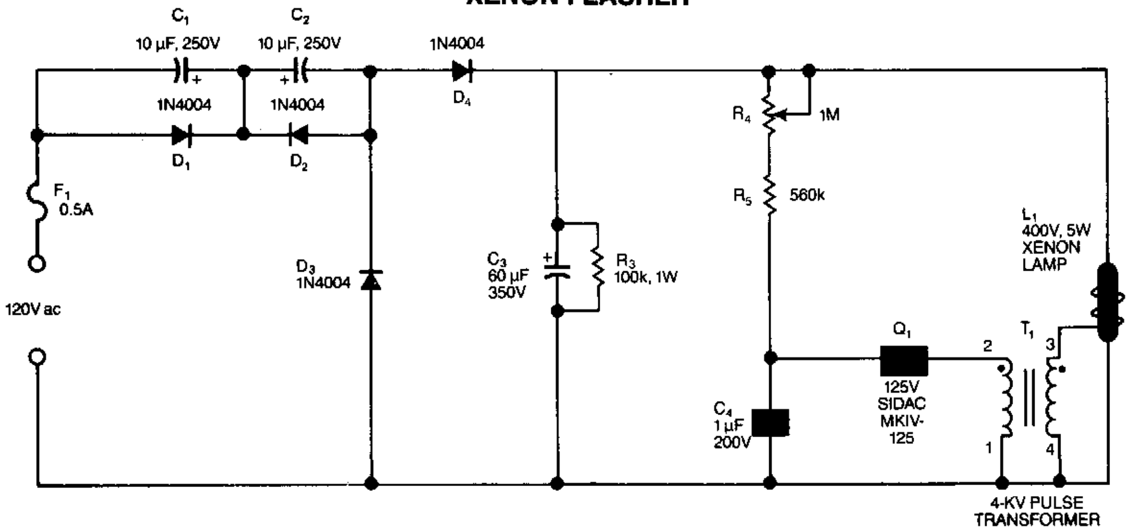
EDN

Fig. 31-1

The pseudorandom sequencer drives a solid-state relay. If you power a low-wattage lamp from the relay, the lamp will appear to flicker like a candle's flame in the wind; using higher-wattage lamps allows you to simulate the blaze of a fireplace or campfire. You can enhance the effect by using three or more such circuits to power an array of lamps.

The circuit comprises an oscillator, IC1, and a 15-stage, pseudorandom sequencer, IC2–4. The sequencer produces a serial bit stream that repeats only every 32 767 bits. Feedback from the sequencer's stages 14 and 15 go through IC4D and back to the serial input of IC2. Notice the RC network that feeds IC4C; the network feeds a positive pulse into the sequencer to ensure that it won't get stuck with all zeros at power-up. The leftover XOR gates IC4A and IC4B further scramble the pattern. The serial stream from IC4B drives a solid-state relay that features zero-voltage switching and can handle loads as high as 1 A at 12 to 280 Vac.

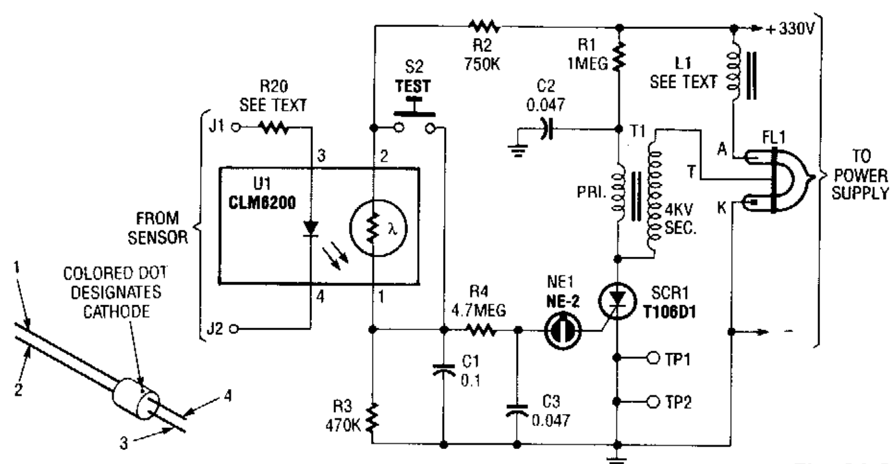
XENON FLASHER



EDN *Fig. 31-2*

Using a voltage-doubler supply, this circuit charges a 60-μF capacitor and discharges it through a Xenon lamp. The SIDAC device is manufactured by Motorola. It is a two-terminal device that breaks over at a specified voltage. R4, R5, and C4 determine the flash rate.

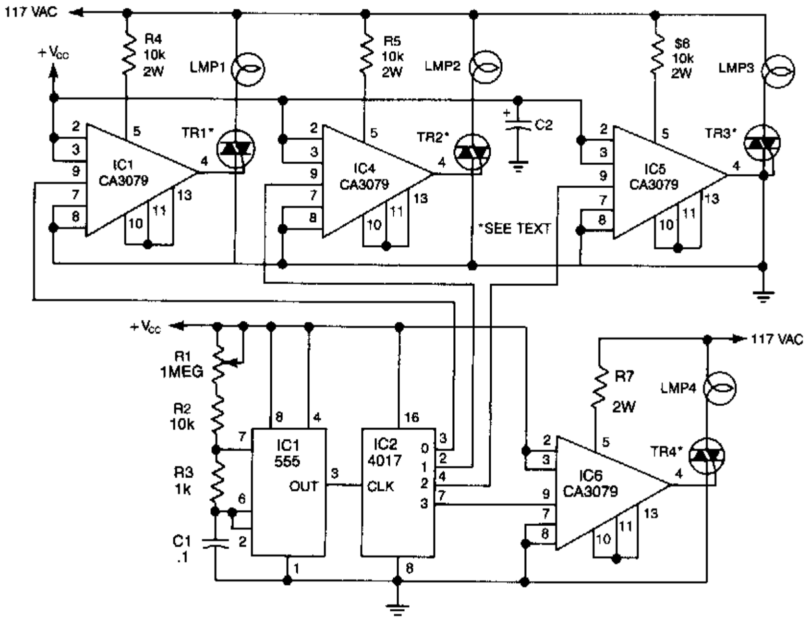
STROBE ALARM



RADIO-ELECTRONICS *Fig. 31-3*

This strobe gives a visual indication of a sensor input. The input signal causes U1, a light dependent resistor, to charge C1 and C3 through R4. When NE1 fires, C3 discharges into SCR1, which triggers it and causes C2 to discharge through trigger transformer T1, which triggers Flashlamp FL1. The 330-V supply should have about 50 to 100 μF output capacitance. L1 supplies about 25-mH inductance to prolong the flash and the life of FL1.

SEQUENTIAL FLASHER



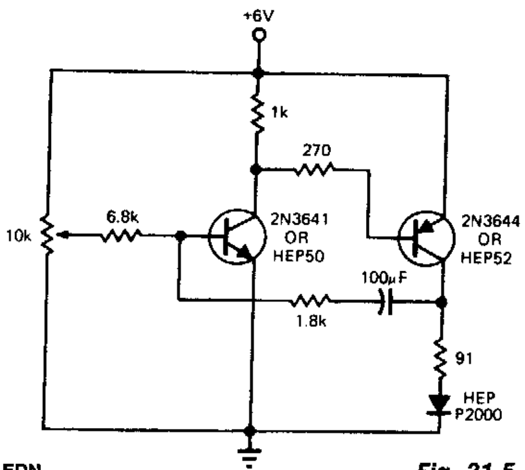
RADIO-ELECTRONICS

Fig. 31-4

Using a 555 timer to drive a CMOS counter, this device uses RCA CA3079 zero-voltage switch to control triacs TR1 through TR4. This circuit can be used to sequence lamp displays, etc.

Caution: The CA3079s are connected to the 117-V line, as is the clock and counter circuit and their power supplies. Use caution, good insulation, and safe construction practices.

LED FLASHER

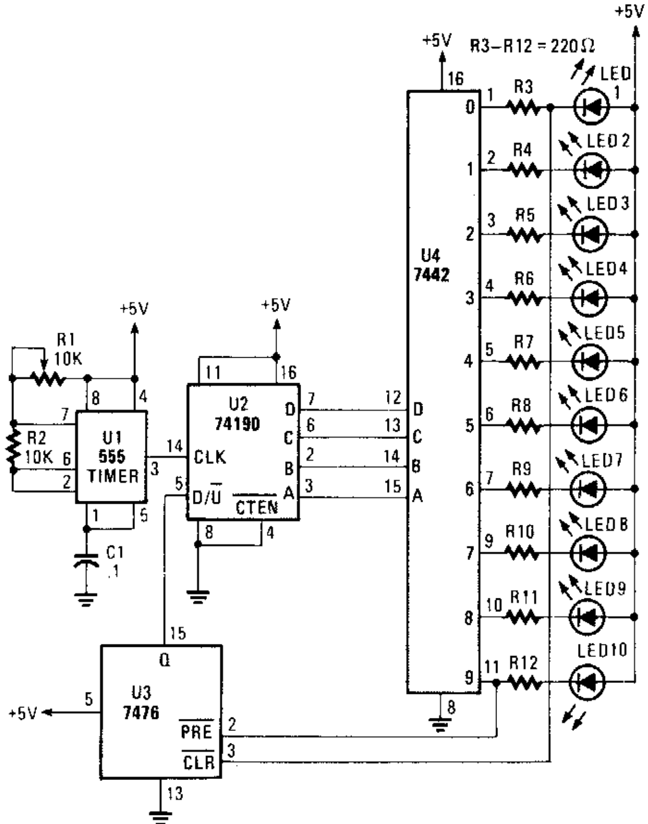


EDN

Fig. 31-5

This circuit is designed to flash an LED. The 100-μF capacitor can be changed to alter the flash rate as desired.

SEQUENTIAL LED FLASHER WITH REVERSIBLE DIRECTION

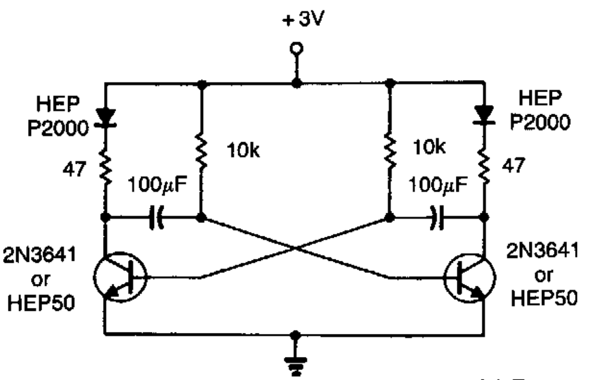


POPULAR ELECTRONICS

Fig. 31-6

A 555 timer clocks a 74190 up/down counter. The 74190 drives BCD decoder driver 7442. The 7476 is used to reverse the count on 0 and 9, which results in an up-down-up-down count sequence.

MULTIVIBRATOR WITH LEDs



A simple astable multivibrator is used to alternately flash two LEDs. The approximate time constant is 0.69.

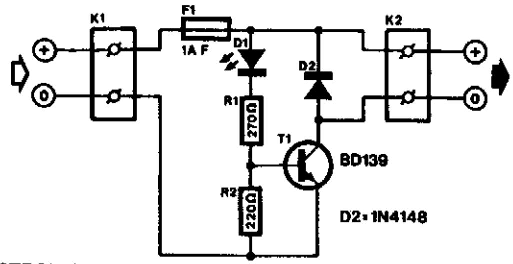
$(R_1C_1 + R_2C_2)$ $R_1 = R_2 = 10\ k\Omega$

$C_1 = C_2 = 100\ \mu F$

EDN

Fig. 31-7

FLASHING LED CONTROLLER



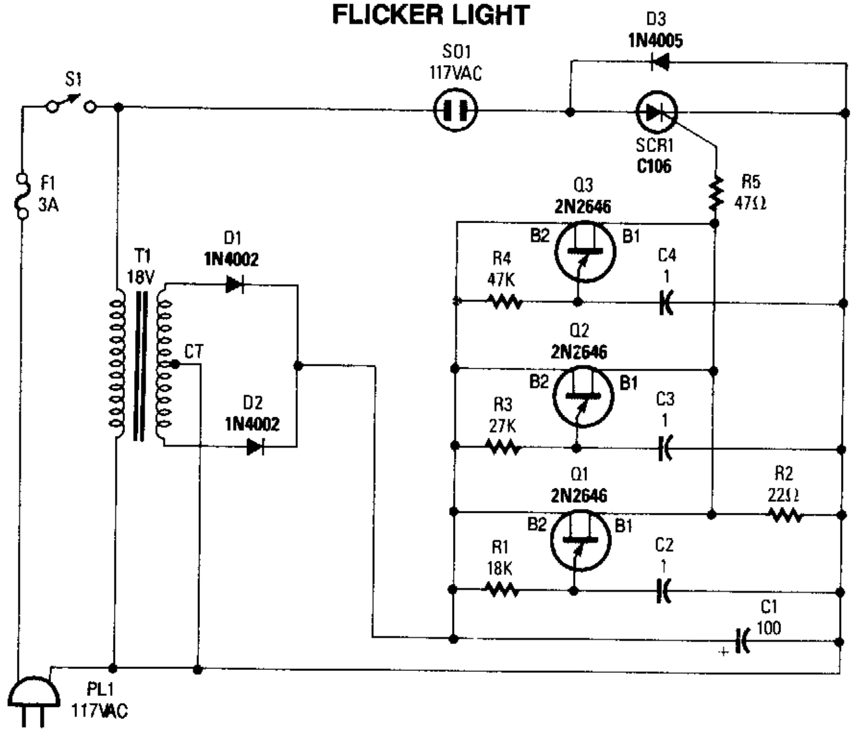
ELEKTOR ELECTRONICS

Fig. 31-8

The LED with integrated flasher is connected in series with the base-emitter junction of transistor T1. Thus, the load connected to K2 is switched on and off in rhythm with the flash rate. This load can be a relay or a lamp.

The maximum collector current of the transistor (of the BD139 = 750 mA) must not be exceeded. If that is not sufficient, a power Darlington can be used, which will give some amperes. The current drawn by the circuit under no-load conditions amounts to 20 mA.

FLICKER LIGHT



POPULAR ELECTRONICS

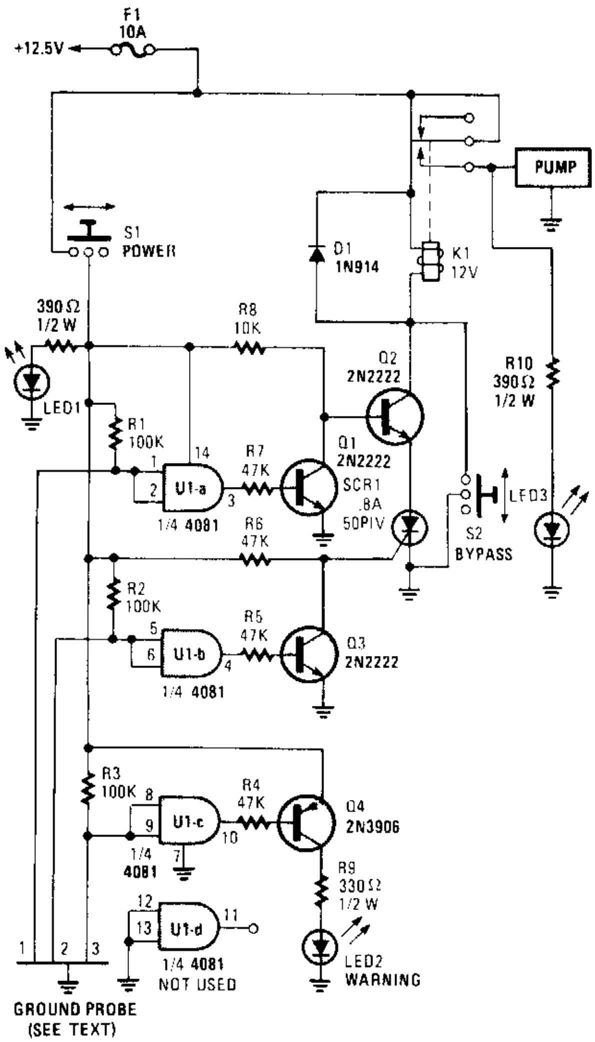
Fig. 31-9

This circuit will produce a flicker light effect with an ordinary incandescent lamp. Three UJT relaxation oscillators fire the SCR in a pattern.

32

Fluid and Moisture Detectors

The sources of the following circuits are contained in the Sources section, which begins on page 666. The figure number in the box of each circuit correlates to the entry in the Sources section.



POPULAR ELECTRONICS

Fig. 32-1

This circuit will power up a water pump when the water reaches a predetermined level. Then it turns itself off when the water recedes to another predetermined point.

Gates U1A through U1C each have their two inputs tied together, and serve as probes. The probes are then placed at various levels to trigger a particular function at a predetermined time. The ground side of the circuit is placed below the minimum water level. The inputs to each gate are tied high through a 100-kΩ resistor connected to the +12.5-V bus.

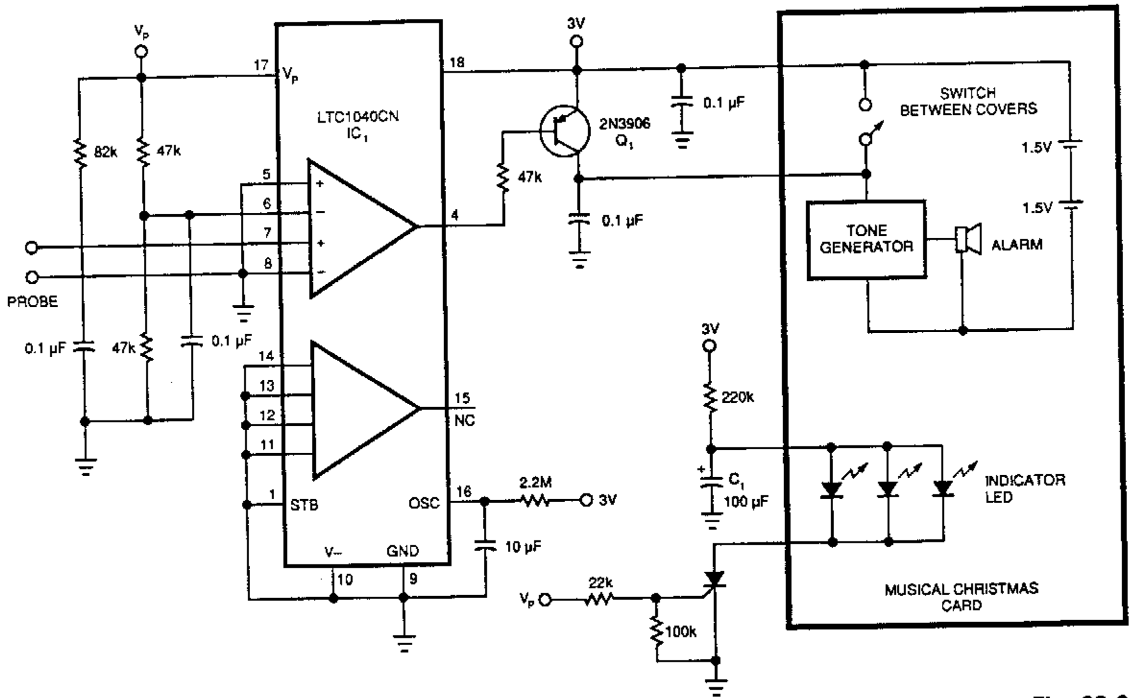
As the water level slowly rises to probe 1, the input to U1A is pulled low by the conduction of current through the water to the ground probe. That turns Q1 off and Q2 on. With Q2 turned on, the circuit is placed in the standby mode, ready to activate the pump when conditions are right.

Probe 2 is placed at the maximum water level. If the water level reaches probe 2, the input of U1B is brought low, turning Q3 on, which, in turn, causes current to be applied to the gate of SCR1, turning it on. The circuit through K1, Q2, and SCR1 is now complete to ground, and the water pump is now turned on, which causes the water level to recede. When the water level falls below probe 2, U1B goes back to logic high.

However, because of the latching nature of SCR1, the pump continues to run until the water level falls below probe 1. At that point, the ground circuit opens and de-energizes K1, which turns the pump off. The pump will not turn on again until the water level again rises above probe 2.

Probe 3 was added as a warning. If the water level reaches probe 3, LED2 indicates that the pump is not working. Switch S2 is a manual override and S1 powers the sensing circuit. LED3 indicates that power has been applied to the pump. LED1 indicates that power has been applied to the sensor.

3-V WATER-LEVEL DETECTOR

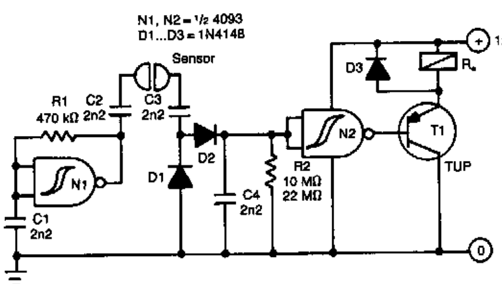


EDN

Fig. 32-2

Originally, this circuit was used to sense a low-water level in a Christmas tree stand, but the circuit can be used as a water-level detector for pump controls, water sensors (for garden and lawn applications), etc. A comparator and probe setup with a Linear Technology LTC1040CN comparator drives a 2N3906, which switches a tone generator. Sampling occurs every 20 seconds, which minimizes current drain. A pair of dry cells will power the circuit for several months.

LIQUID-LEVEL SENSOR

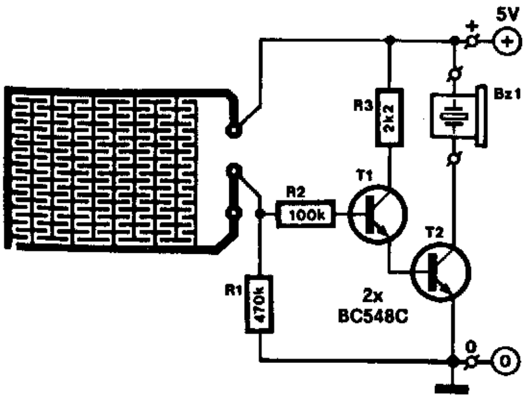


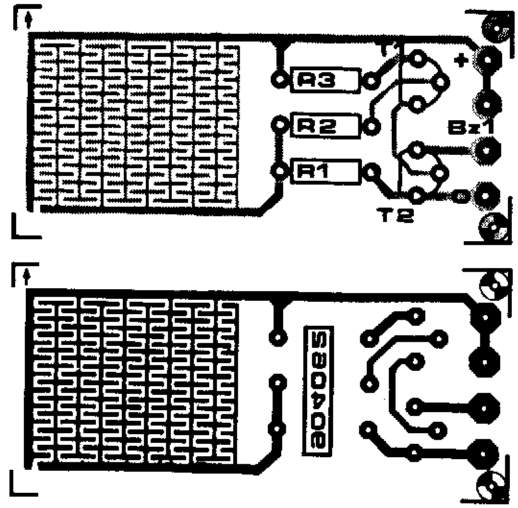
This circuit uses an ac-sensing signal to eliminate electrolytic corrosion. The ac signal is rectified and used to drive a transistor that controls a relay.

ELEKTOR ELECTRONICS

Fig. 32-3

FULL BATHTUB INDICATOR





ELEKTOR ELECTRONICS

Fig. 32-4

Running a bath can end in a minor domestic disaster if you forget to turn off the taps in time. This indicator activates an active buzzer to provide an audible warning when a given water level is reached.

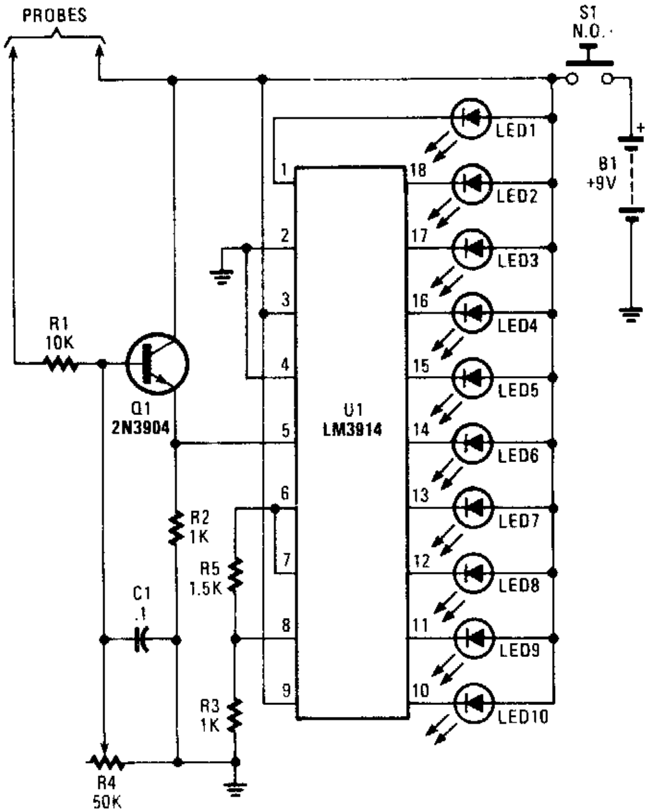
Because the water sensor and the driver circuit for the buzzer are contained on one PC board, the indicator, together with the 9-V battery and the buzzer, can be built into a compact case. Obviously, the sensor, which is etched on the PC-board, must not be fitted in case-iron or steel bath, the indicator is secured to it with the aid of a magnet glued onto the case. To prevent scratching the bath, the magnet can be covered in plastic or rubber. If you have a polypropylene bath, the indicator can be stuck to it with blue tack or double-sided adhesive tape.

When the water reaches the sensor, the base of T1 is connected to the positive supply line. As a result, T1 and T2 are switched on so that the buzzer BZ1, a self-oscillating type, is activated. The current drawn by the circuit in that condition is about 25 mA.

In case the circuit is actuated by steam, its sensitivity can be reduced by increasing the value of R2. It is best to tin the PC board tracks to prevent corrosion.

T1, T2 = BC548C
BZ1 = active piezo-ceramic resonator

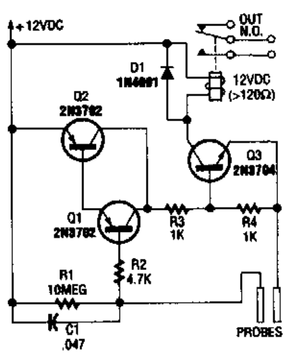
MOISTURE DETECTOR



POPULAR ELECTRONICS *Fig. 32-5*

A bar-graph LED driver is used to drive 10 LEDs to give a relative indication of moisture. The moisture probes are connected so that electrical conductivity due to moisture tends to forward bias Q1, providing a dc voltage at pin 5 of U1 that is proportional to leakage current. Ideally, the probes should be made of stainless steel.

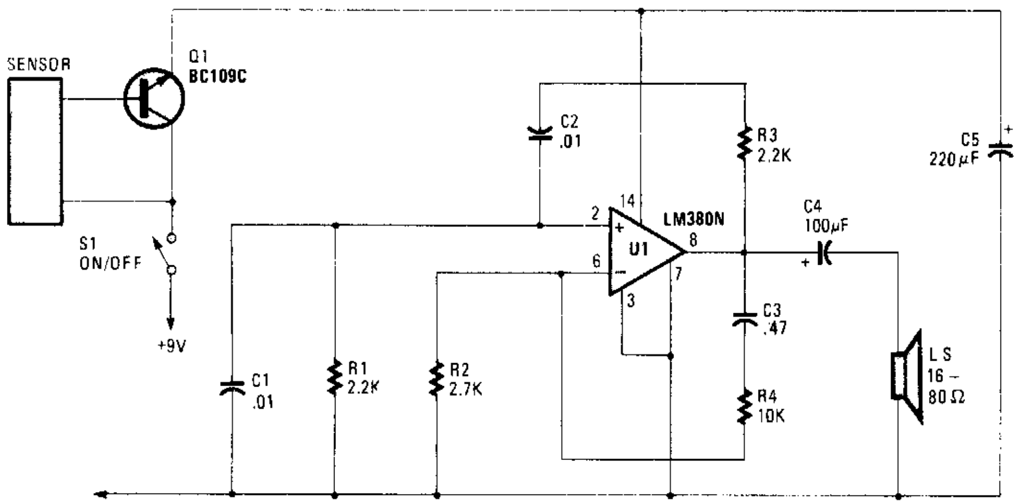
FLOOD ALARM



RADIO-ELECTRONICS *Fig. 32-6*

Using a few bipolar transistors, this circuit acts as a flood alarm. When liquid touches the probes, leakage current biases Q1, Q2, and Q3 (a dc-coupled amplifier) into conduction, which activates the relay. The contacts can be hooked into the alarm system.

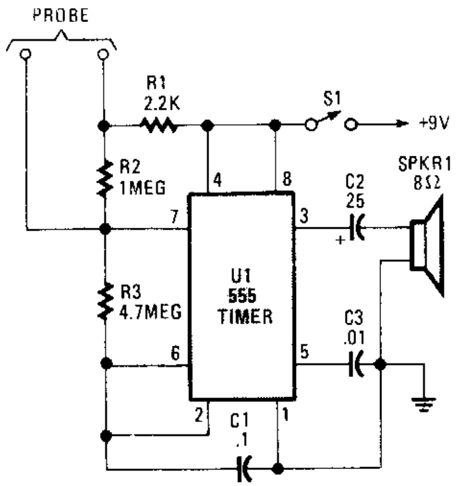
RAIN ALARM



POPULAR ELECTRONICS *Fig. 32-7*

This rain sensor causes Q1 to conduct when conductive liquid (rainwater, etc.) applies bias to its base. This bias triggers LM380N oscillator and causes LS to emit a tone.

FULL-CUP DETECTOR FOR THE BLIND

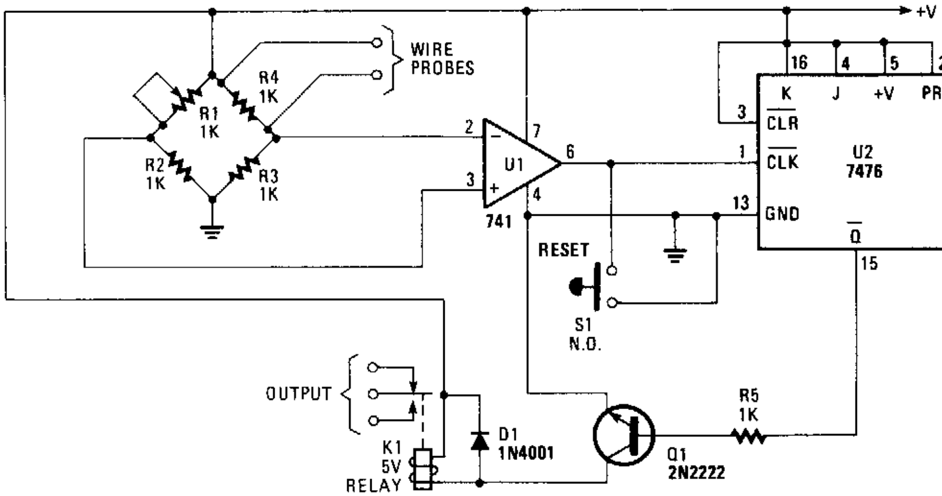


POPULAR ELECTRONICS *Fig. 32-8*

At the heart of the Full-Cup Detector is a 555-oscillator/timer configured to produce a 15-Hz click, until its probe contacts are bridged, at which time its output frequency goes to about 500 Hz.

This circuit can be used by the visually handicapped to determine when a cup or bowl is full of liquid (coffee, soup, etc). U1, an NE555, produces ticks at 15 Hz. A set of probes (wire, etc.) is placed in the container at the desired level. When the liquid level contacts the probes, the frequency of clicks increases to several hundred hertz, depending on its conductivity.

LATCHING WATER SENSOR

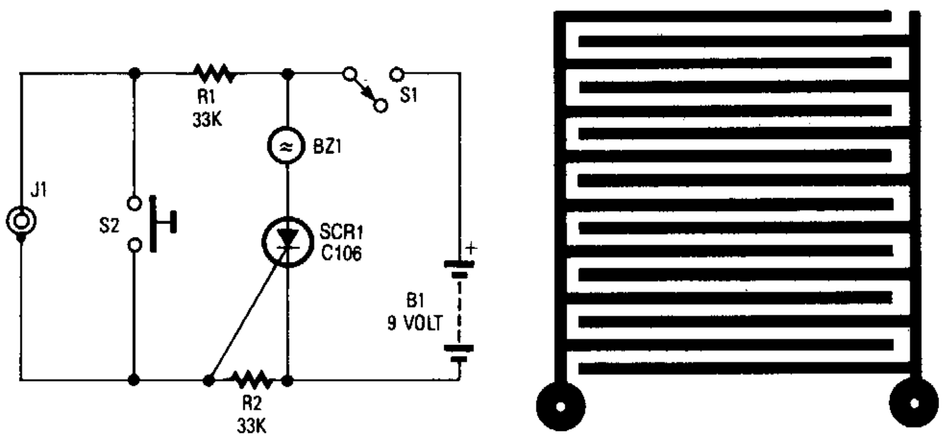


POPULAR ELECTRONICS

Fig. 32-9

A balanced Wheatstone bridge controls a JK flip-flop that uses an op amp as an interface. This in turn drives a relay circuit. R1 through R4 can be made larger for increased sensitivity.

WATER-LEAK ALARM



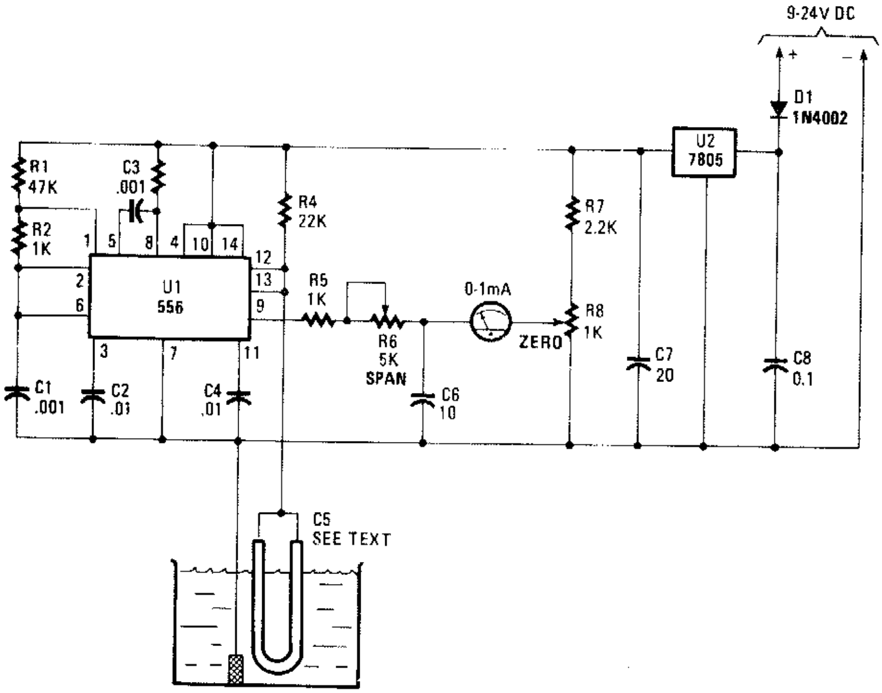
POPULAR ELECTRONICS

If you choose to make your own moisture sensor, this foil pattern should come in handy.

Fig. 32-10

A sensor connected to J1 causes SCR1 to conduct, which sounds buzzer BZ1. The sensor is a PC-board foil pattern grid. Several sensors can be wired in parallel for increased coverage or to monitor several places simultaneously.

WATER-LEVEL MEASUREMENT CIRCUIT



POPULAR ELECTRONICS *Fig. 32-11*

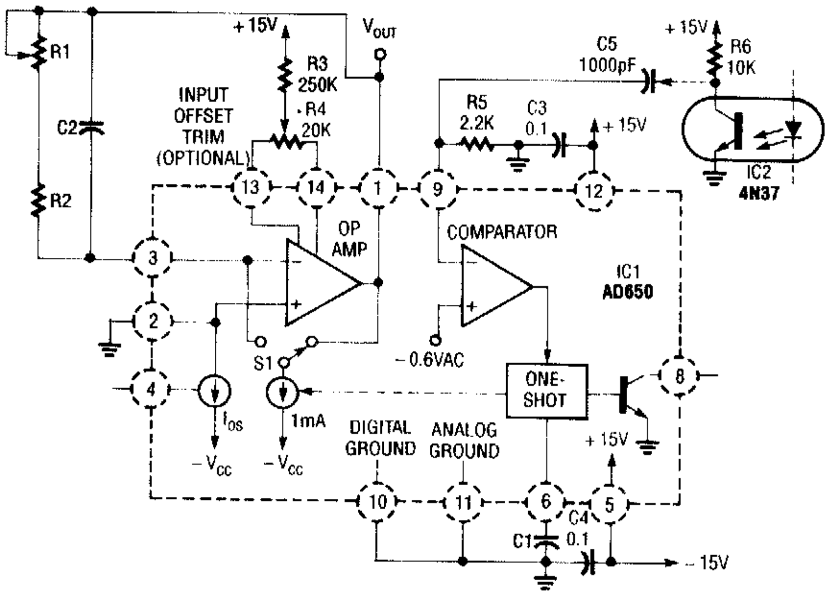
Using a capacitor sensor to detect a water level is a simple method of sensing. This circuit uses C5, which is 10″ to 20″ of #22 enamelled wire as one electrode. This shifts the oscillator, an NE556 timer, in frequency. The frequency shift depends on the capacitance charge, which in turn varies with water level. A meter connected to pin 9 of the 556 is used as an indicator. C5 can be made larger or smaller to suit the intended application.

33

Frequency-to-Voltage Converters

The sources of the following circuits are contained in the Sources section, which begins on page 666. The figure number in the box of each circuit correlates to the entry in the Sources section.

FREQUENCY/VOLTAGE CONVERTER WITH OPTOCOUPLER INPUT



CIRCUIT VALUES

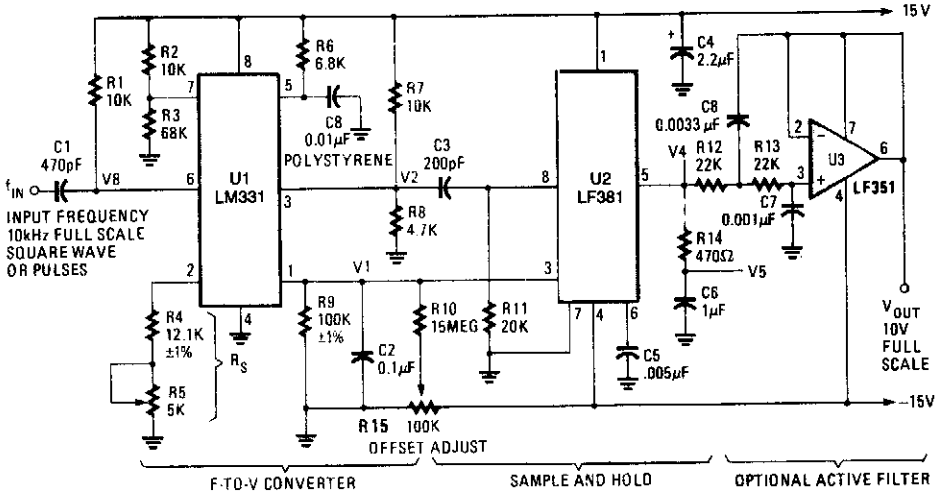
Full-Scale Frequency	Full-Scale Output	C1	C2	R1 (ohms)	R2 (ohms)
10 kHz	1 V	3300 pF	3.3 μF	1K	3.8K
10 kHz	10 V	3300 pF	0.33 μF	10K	38.3K
100 kHz	1 V	680 pF	0.33 μF	500	1.82K
100 kHz	10 V	680 pF	3300 pF	5K	18.2K
1 MHz	1 V	47 pF	3300 pF	500	1.33K
1 MHz	10 V	47 pF	1000 pF	5K	13.3K

RADIO-ELECTRONICS

Fig. 33-1

In this circuit, the input from IC2 optocoupler is fed to the comparator input of the AD650 (Analog Devices or Maxim Electronics) V/F converter. This internally generates a pulse that is fed to the op amp, which outputs a dc voltage that is proportional to frequency. Component values are shown in the figure.

FREQUENCY/VOLTAGE CONVERTER WITH SAMPLE AND HOLD



POPULAR ELECTRONICS

Fig. 33-2

U1 is a frequency/voltage converter, feeding sample-and-hold circuit using an LF381. An LF351 provides 10-V full-scale output. The circuit produces 1-V/kHz output.

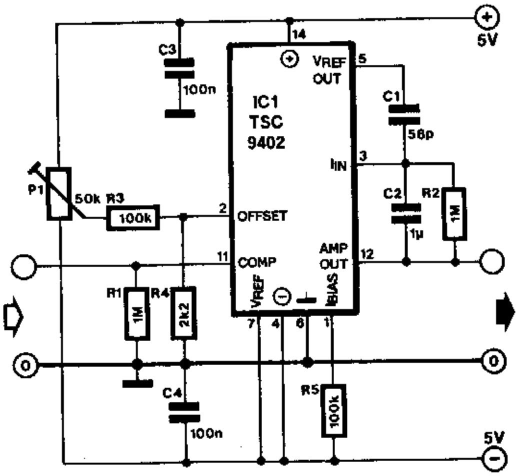
FREQUENCY/VOLTAGE CONVERTER

Teledyne Semiconductor's Type TSC9402 is a versatile IC. Not only can it convert voltage into frequency, but also frequency into voltage. It is thus eminently suitable for use in an add-on unit for measuring frequencies with a multimeter. Only a few additional components are required for this.

Just one calibration point sets the center of the measuring range (or of that part of the range that is used most frequently). The frequency-proportional direct voltage at the output (pin 12—AMP OUT) contains interference pulses at levels up to 0.7 V. If these have an adverse effect on the multimeter, they can be suppressed with the aid of a simple RC network. The output voltage, U_o, is calculated by:

$$U_o = U_{ref}(C_1 + 12\ \text{pF})\ R_2 f_{in}$$

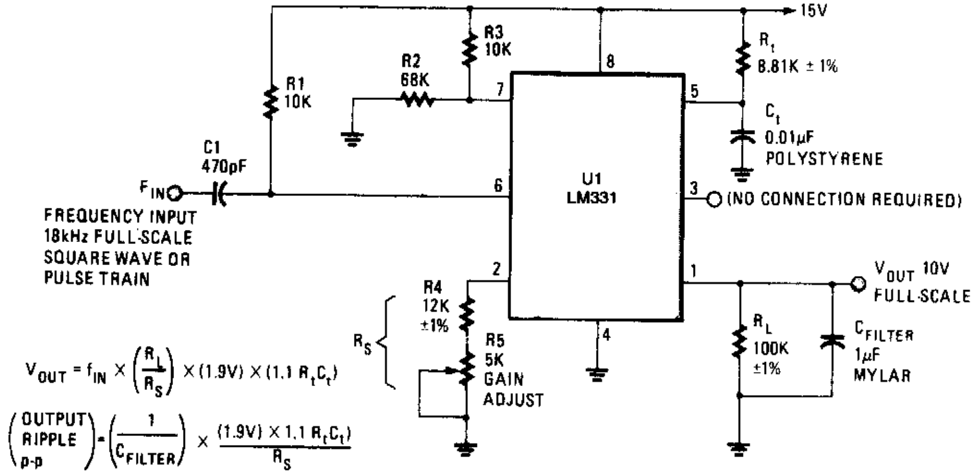
Because the internal capacitance often has a greater value than the 12 pF taken here, the formula does not yield an absolute value. The circuit has a frequency range of dc to 10 kHz. At 10 kHz, the formula gives a value of 3.4 V. The circuit draws a current of not more than 1 mA.



ELEKTOR ELECTRONICS

Fig. 33-3

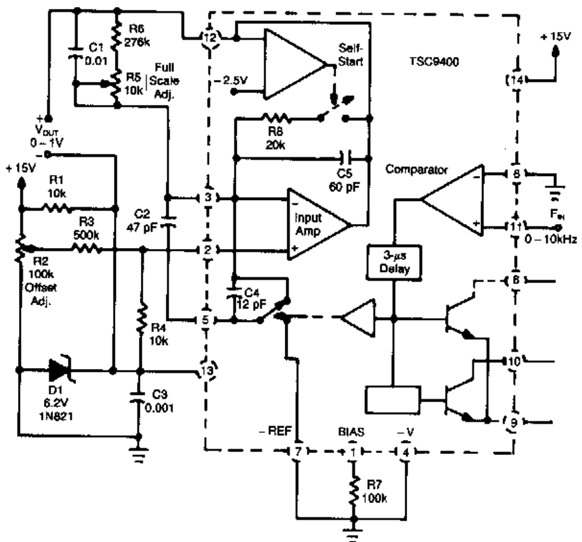
FREQUENCY/VOLTAGE CONVERTER



POPULAR ELECTRONICS *Fig. 33-4*

A dc output that is proportional to frequency can be derived with this circuit. It is useful for analog frequency meter or tachometer applications.

SINGLE-SUPPLY FREQUENCY/VOLTAGE CONVERTER



RADIO-ELECTRONICS *Fig. 33-5*

A Teledyne TSC9400 provides 0-to-1-V output from a 0-to-10-kHz input. A single +15-V supply is used. Linearity is 0.25% to 10 kHz.

34

Function Generators

The sources of the following circuits are contained in the Sources section, which begins on page 667. The figure number in the box of each circuit correlates to the entry in the Sources section.

AUDIO FUNCTION GENERATOR

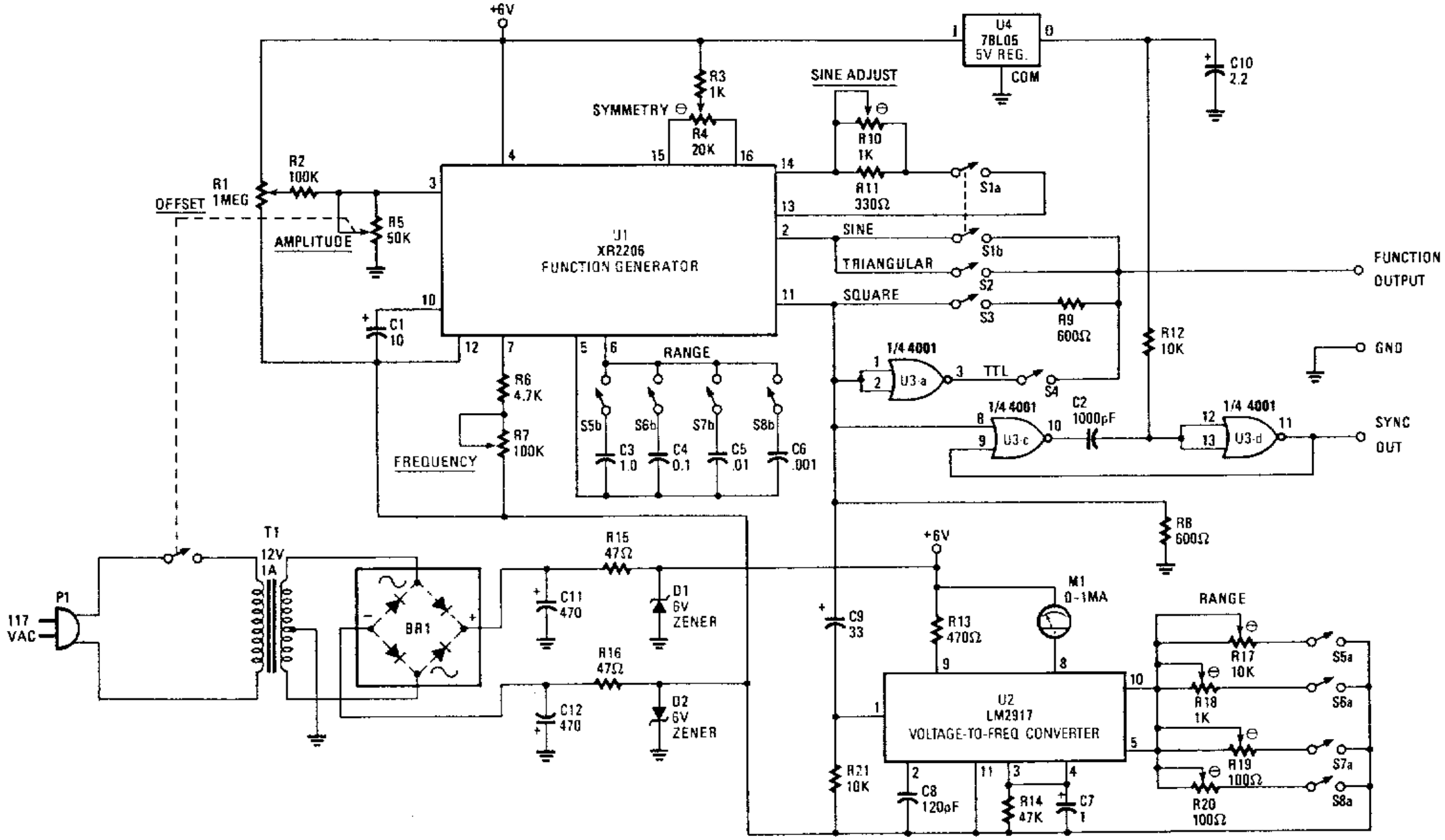
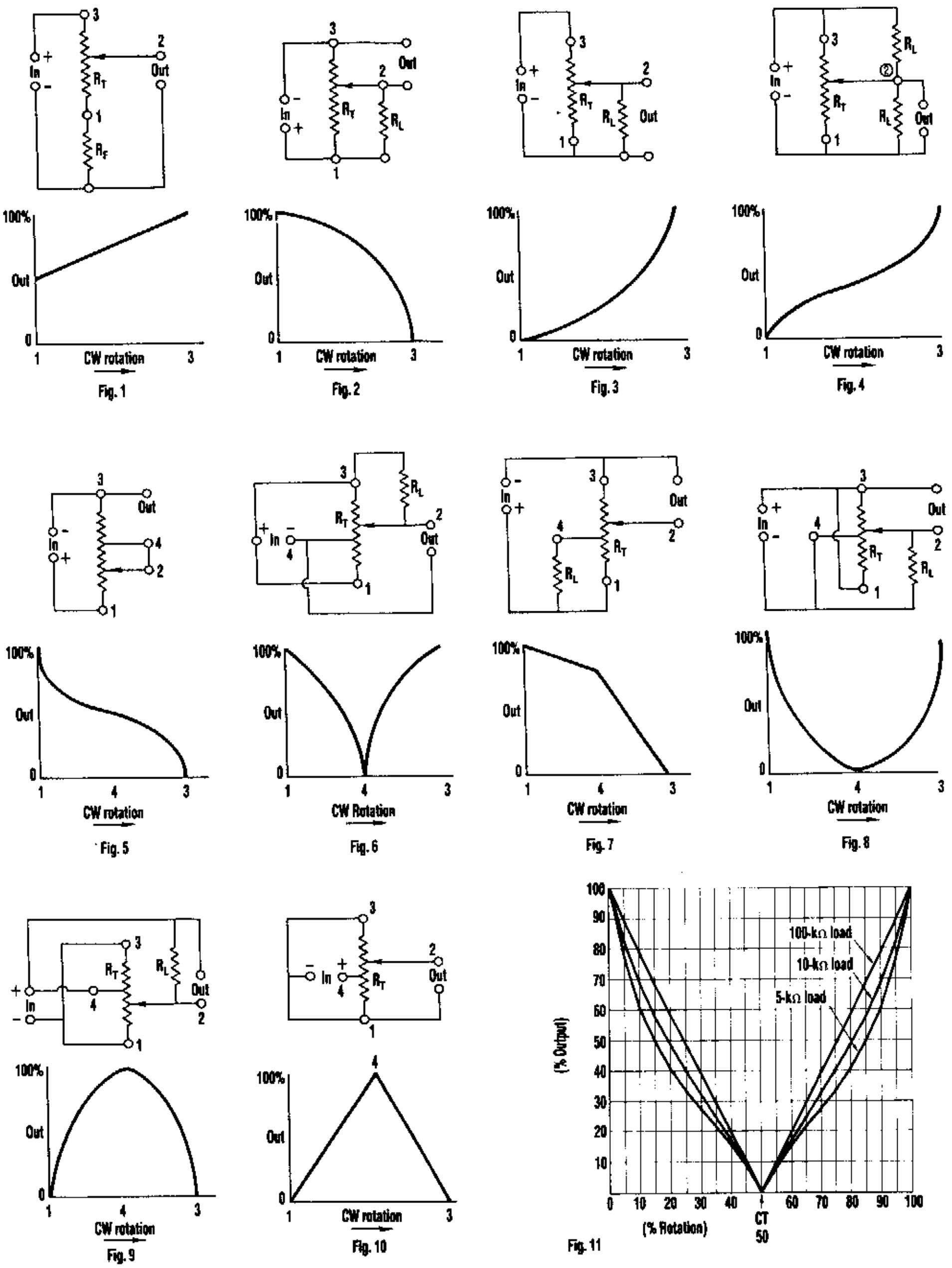


Fig. 34-1

Using an EXAR XR2206, this generator will produce sine, square, and triangular waves from 10 Hz to 100 kHz. U1 is the XR2206 chip, R7 controls frequency, and S5 through S8 select the frequency range. U3 produces a TTL-compatible square-wave output, while U3C and D produce a sync signal for scope use. U2 is a frequency/voltage converter that is used to drive analog meter M1, which reads the generator frequency.

NONLINEAR POTENTIOMETER OUTPUTS

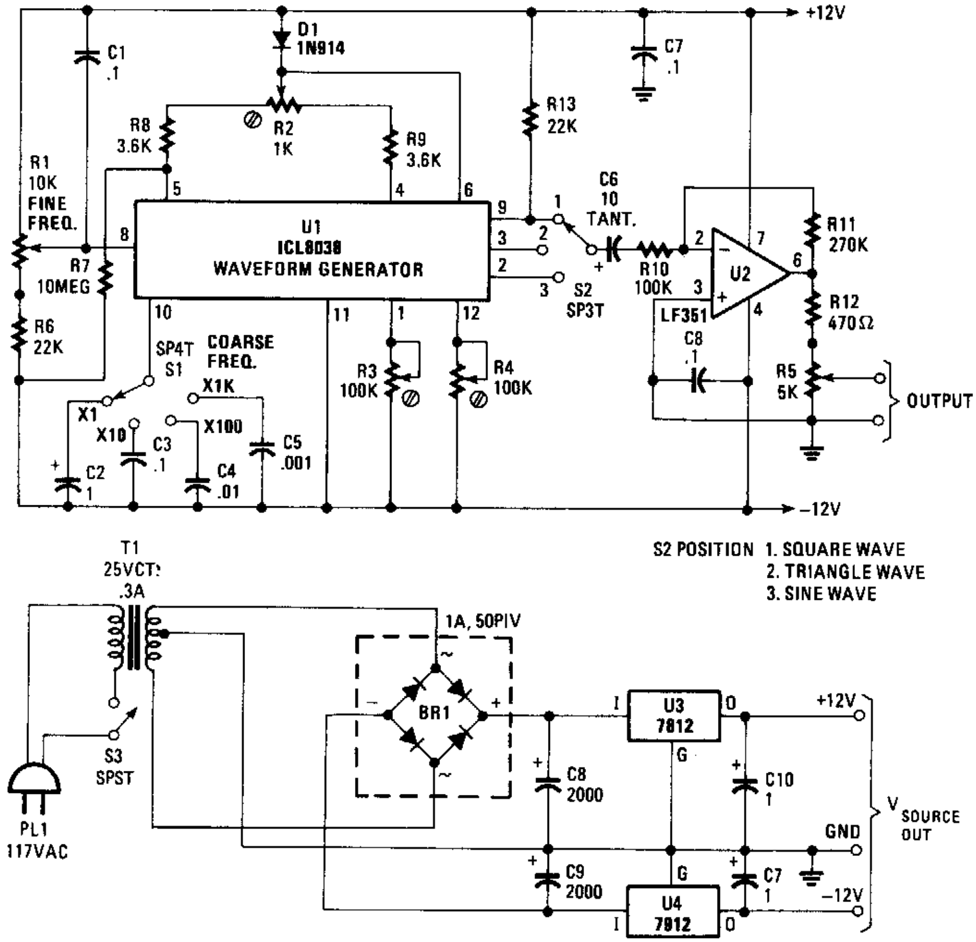


ELECTRONIC DESIGN

Fig. 34-2

Using these illustrated configurations, various rotation output characteristics can be obtained from potentiometers and one or two resistors.

FUNCTION GENERATOR

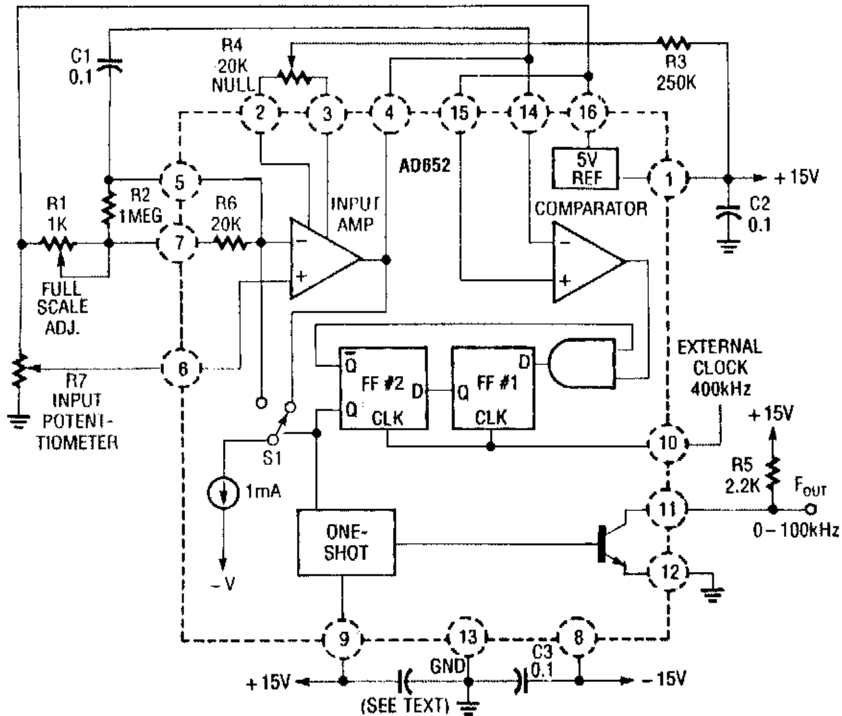


RADIO-ELECTRONICS

Fig. 34-3

Using an Intersil ICL8038, this function generator generates frequencies from 1 Hz to over 80 kHz. R1 is the fine frequency control and S1 is the coarse frequency control range switch. S2 selects square-, triangle-, or sine-wave output. U2 is a buffer amplifier and R5 sets output level. R2 is adjusted for a symmetrical triangle wave. R3 and R4 are adjusted for minimum sine-wave distortion. Power supply is ±12 V at less than 100 mA.

POTENTIOMETER-POSITION V/F CONVERTER

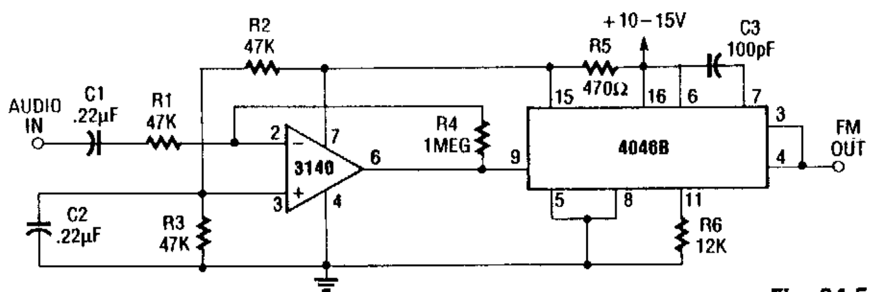


RADIO-ELECTRONICS

Fig. 34-4

In this application, an AD652IC is used in a synchronized V/F converter that derives its input from the position of a potentiometer. This can represent a position of a mechanical component, weight, size, etc., to give a 0-to-100-kHz output versus the 0-to-5-V output from the potentiometer.

FM GENERATOR

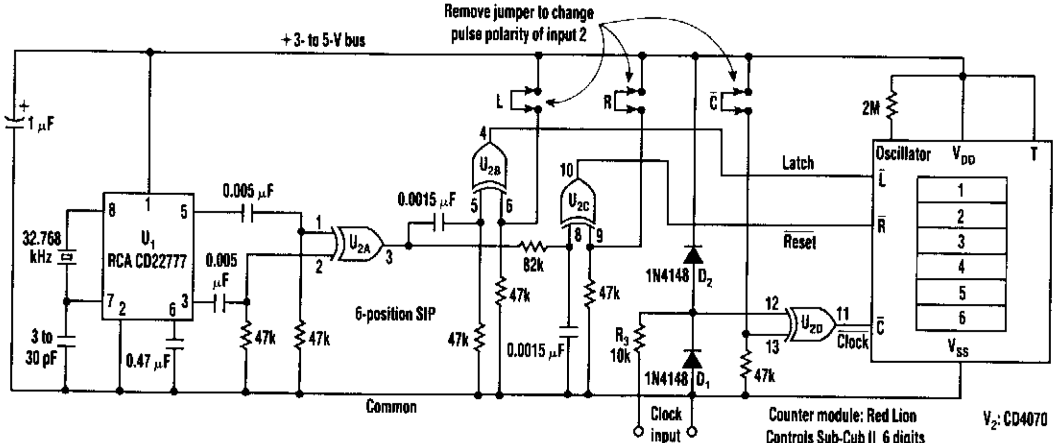


RADIO-ELECTRONICS

Fig. 34-5

The internal zener on pin 15 of the 4046B supplies a stable voltage to the 3140IC op amp. This amplifier modulates the 4046B VCO. The amplifier gain is about 20× (26 dB voltage). The VCO produces a 220-kHz carrier that is FM modulated. C3 can be changed to vary this frequency.

1-Hz TIMEBASE FOR READOUT AND COUNTER APPLICATIONS



ELECTRONIC DESIGN

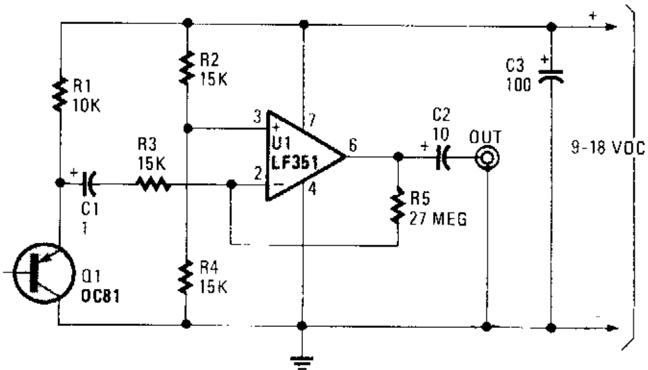
Fig. 34-6

This counter makes direct readout of frequency-generating equipment very easy when a 1-Hz time-base is added to latch, reset, and the count signal is conditioned. This design has the flexibility to select either polarity.

By differentiating, inverting, and ORing the clock pulses in XOR gate U2A, a stream of 1-Hz, positive, 200-μs pulses is generated. For latching, the 1-Hz stream is again differentiated in U2B, input 1 to supply a 50-μs pulse. Though U2B's output goes from high to low, it can be reversed, by making input 2 low.

Because the reset pulse must occur after the latch signal, the 1-Hz stream from U2A is delayed 100 μs at U2C input 1. The output-pulse polarity is determined by making U2C's input 2 either high or low.

WHITE NOISE GENERATOR

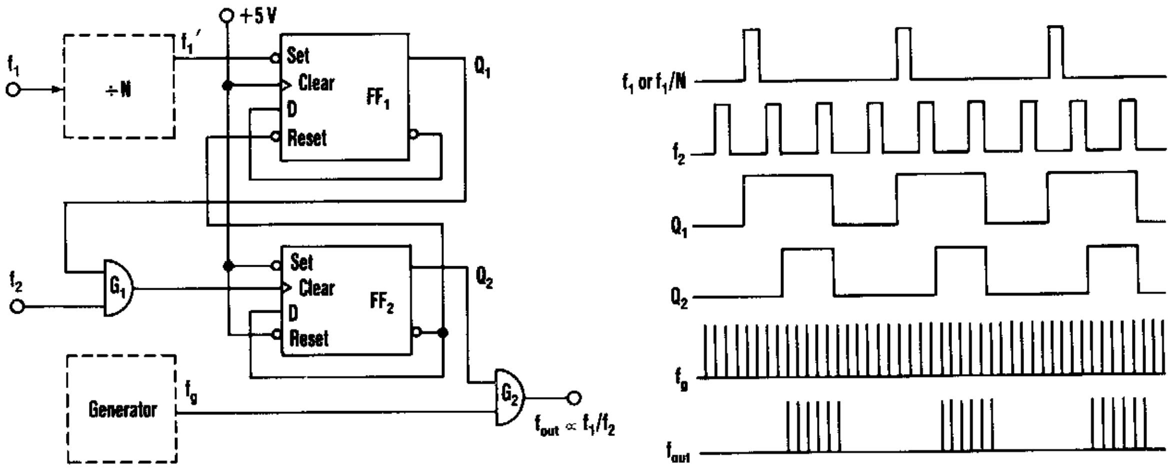


HANDS-ON ELECTRONICS

Fig. 34-7

Germanium transistor Q1 is used as a noise generator in the audio range. U1 acts as a high-gain amplifier. Q1 is not critical; most germanium transistors appear to be satisfactory. A germanium diode can also be substituted. This circuit is mainly used for sound effects and noise experiments. It is not flat over the audio range because of unpredictable effects in Q1, but it should be useful where high precision is not necessary.

FREQUENCY-RATIO MONITORING CIRCUIT

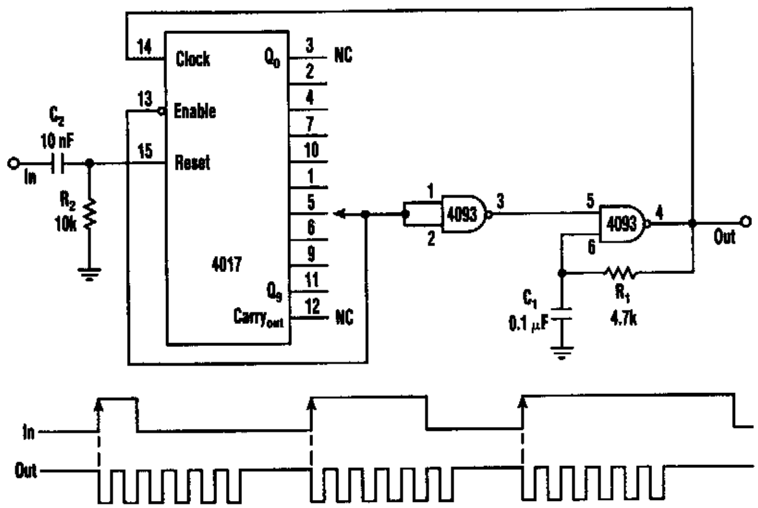


ELECTRONIC DESIGN

Fig. 34-8

This circuit produces an output frequency that is linearly proportional to the ratio of two input frequencies f_1/f_2. Each pulse of the bias f_1 (or f) will open G1 for a period $T = 1/f_2$ so that f_g/f_2 pulses pass to the output.

PULSE TRAIN



ELECTRONIC DESIGN

Fig. 34-9

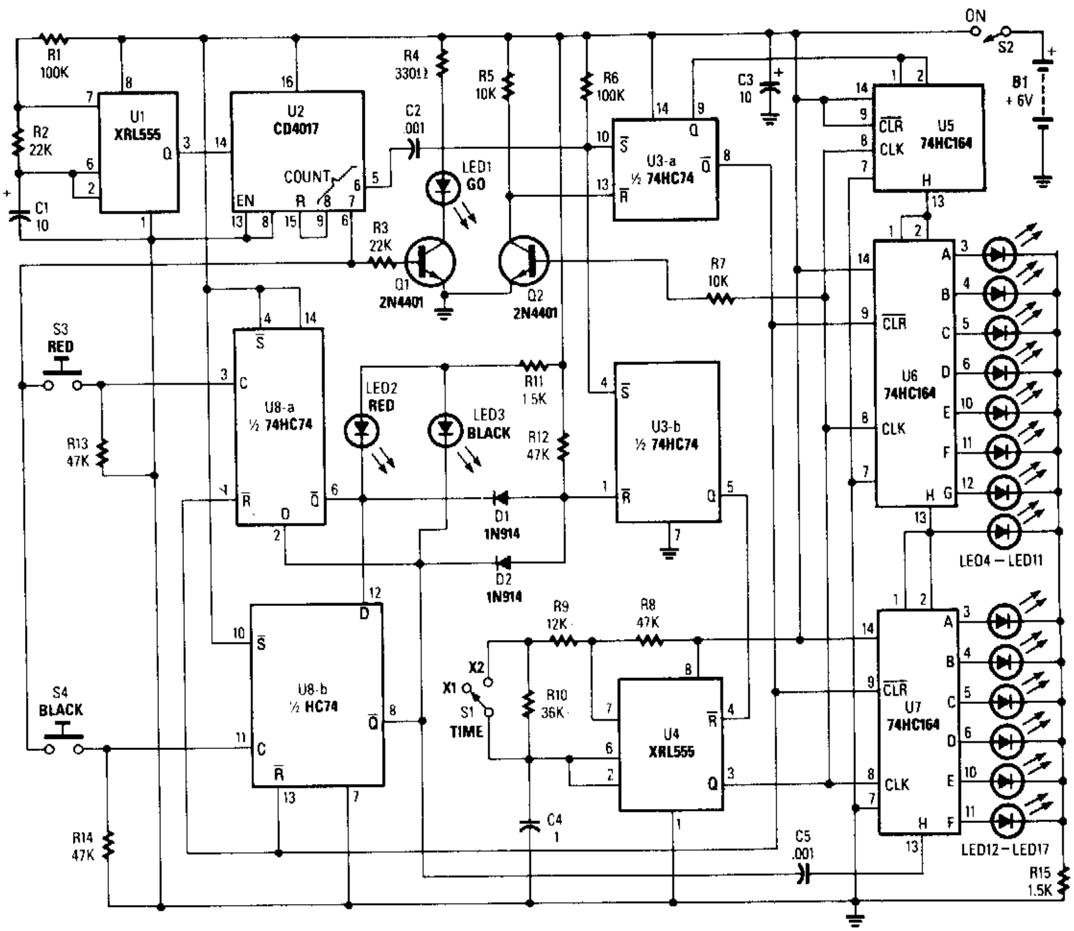
This circuit has a rate multiplier using a 4093 Schmitt trigger as an oscillator, driving a 4017 decade counter. When a pulse present at the input (to C2) 4017 is reset, output zero goes high, and outputs 1 to 9 go low. The oscillator (4093) starts running and the 4017 counts the pulses until the 4017 output (1 to 9) connected to pin 1 and 2 of the 4093 goes high. The oscillator is inhibited and the output remains high until the next input pulse.

35

Games

The sources of the following circuits are contained in the Sources section, which begins on page 667. The figure number in the box of each circuit correlates to the entry in the Sources section.

REACTION TIMER

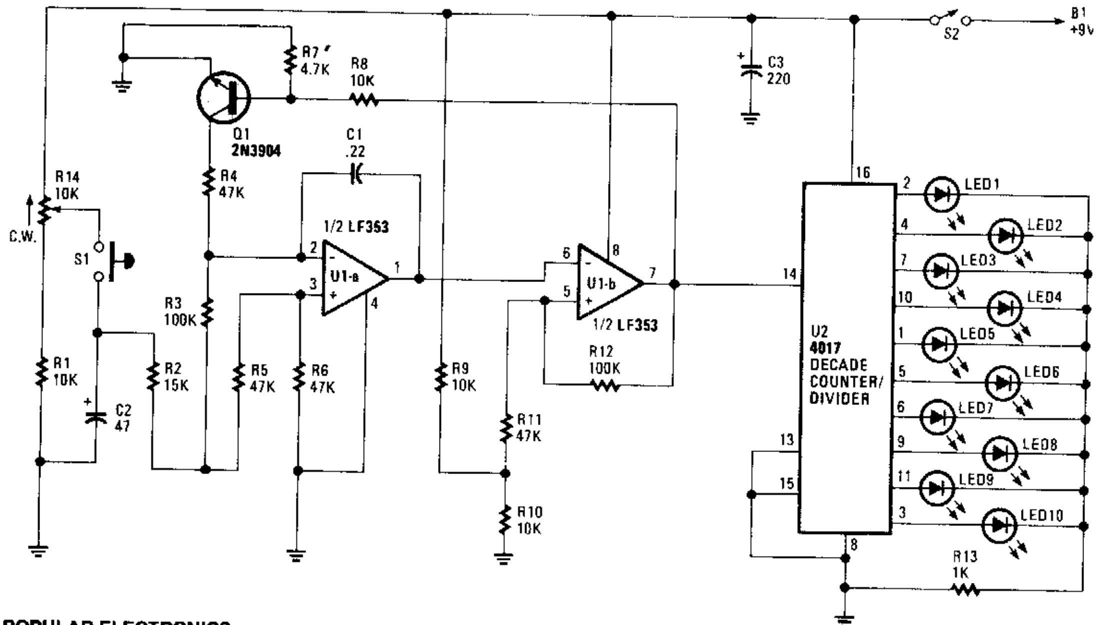


RADIO-ELECTRONICS

Fig. 35-1

This circuit uses a timer to generate pulses at a 5-ms clock rate. The pulses are shifted into the shift register, one at a time, lighting an LED. An auxiliary timer that generates one pulse per second is used to generate timing to activate the "go" LED and start the 5 ms pulses clocking into the registers. At the GO signal each player presses his buttons (S3 or S4). The delay (reaction time) is read out on LED 4 to LED 17; after six seconds, the sequence repeats.

ELECTRONIC ROULETTE GAME

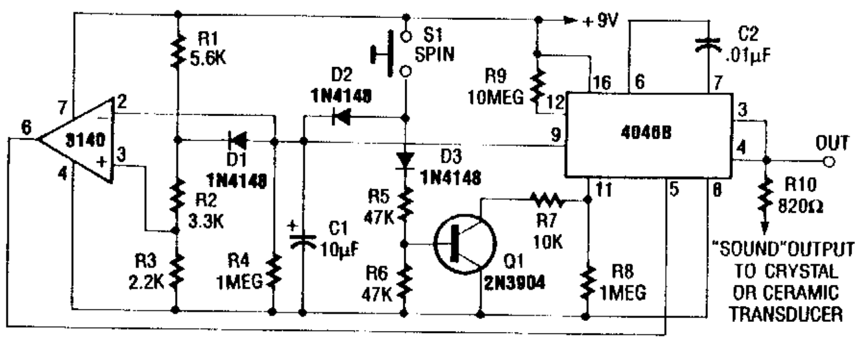


POPULAR ELECTRONICS

Fig. 35-2

R14 is set for an initial "starting" speed of the oscillator U1A and U1B. As C2 charges, oscillation begins slowing down as C2 discharges, giving a roulette-wheel effect on LED S1 through 10. The LED that remains on is the winning number.

RUN-DOWN CLOCK/SOUND GENERATOR

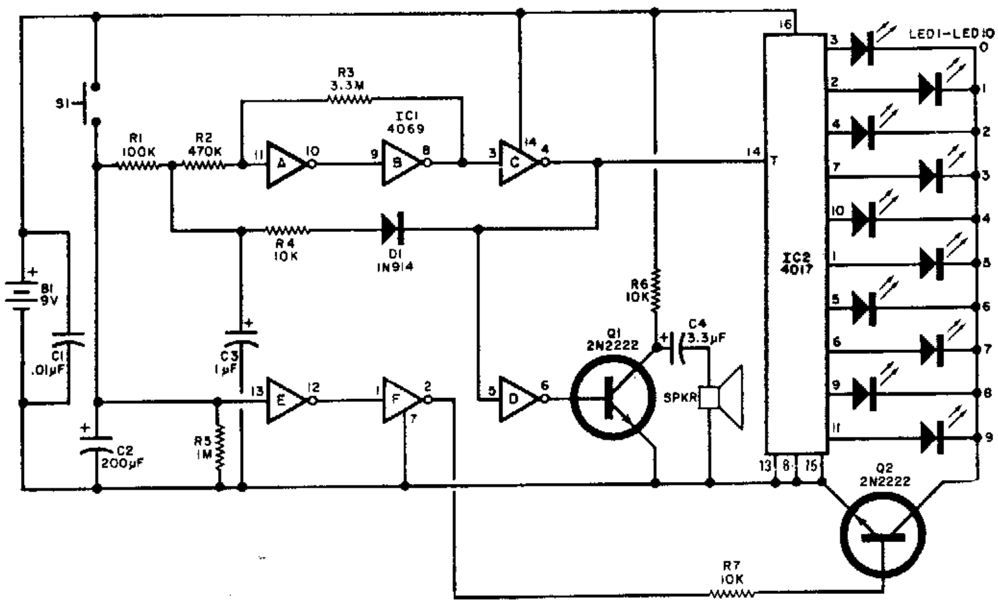


RADIO-ELECTRONICS

Fig. 35-3

Used in electronic roulette or dice games, this circuit produces a clock signal that initially is several tens of kHz (depending on C2) and gradually decreases to zero in about 15 seconds, as C1 discharges through R4.

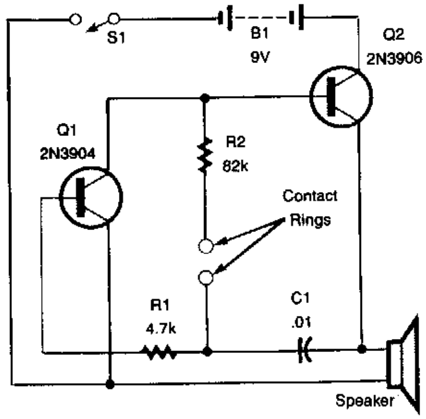
WHEEL OF FORTUNE



POPULAR ELECTRONICS *Fig. 35-4*

This circuit is a 10-LED spinning wheel that "clicks" as the wheel passes each point. The rotation starts fast, then gradually slows down to a random stop (with a click at each position). After the rotation ceases, the selected LED stays lit for about 10 seconds, then goes out. The cycle restarts by depressing the pushbutton switch.

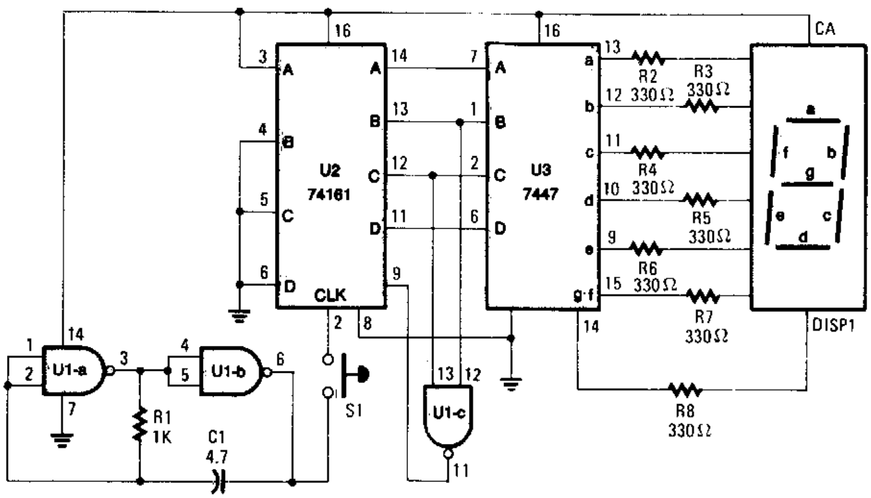
SIMPLE LIE DETECTOR



POPULAR ELECTRONICS *Fig. 35-5*

The variation in skin resistance of the subject is used to vary the frequency of a tone oscillator. The contact rings are two brass rings, about 3/4" ID.

ELECTRONIC DICE



POPULAR ELECTRONICS *Fig. 35-6*

When S1 is pressed, counter U2 is driven by oscillator U1A/U1B and the count (0 through 6) is read on DISP1. R1 and C1 determine the count rate, which should be fast enough to ensure a "random" count.

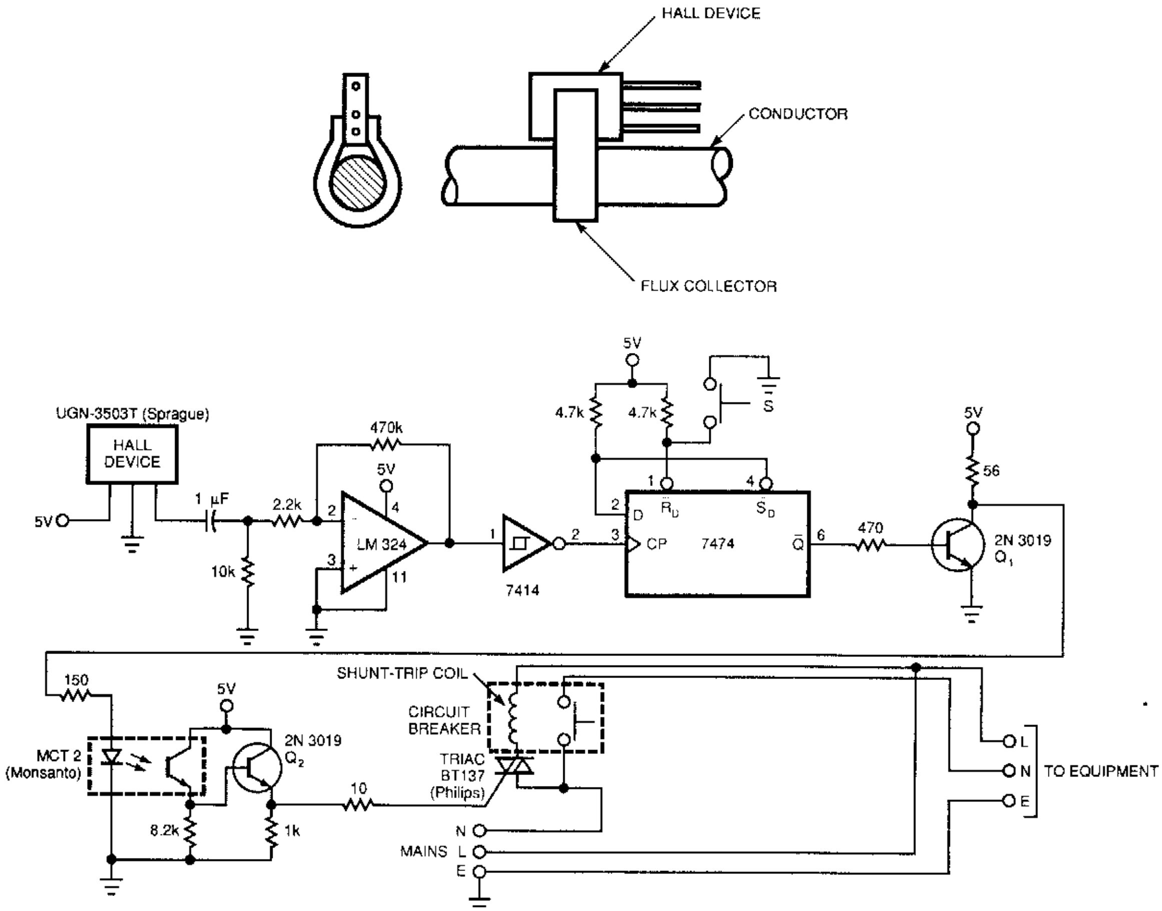
36

Ground-Fault Hall Detector

The source of the following circuit is contained in the Sources section, which begins on page 667. The figure number in the box correlates to the entry in the Sources section.

Ground-Fault Hall Sensor

GROUND-FAULT HALL SENSOR



EDN *Fig. 36-1*

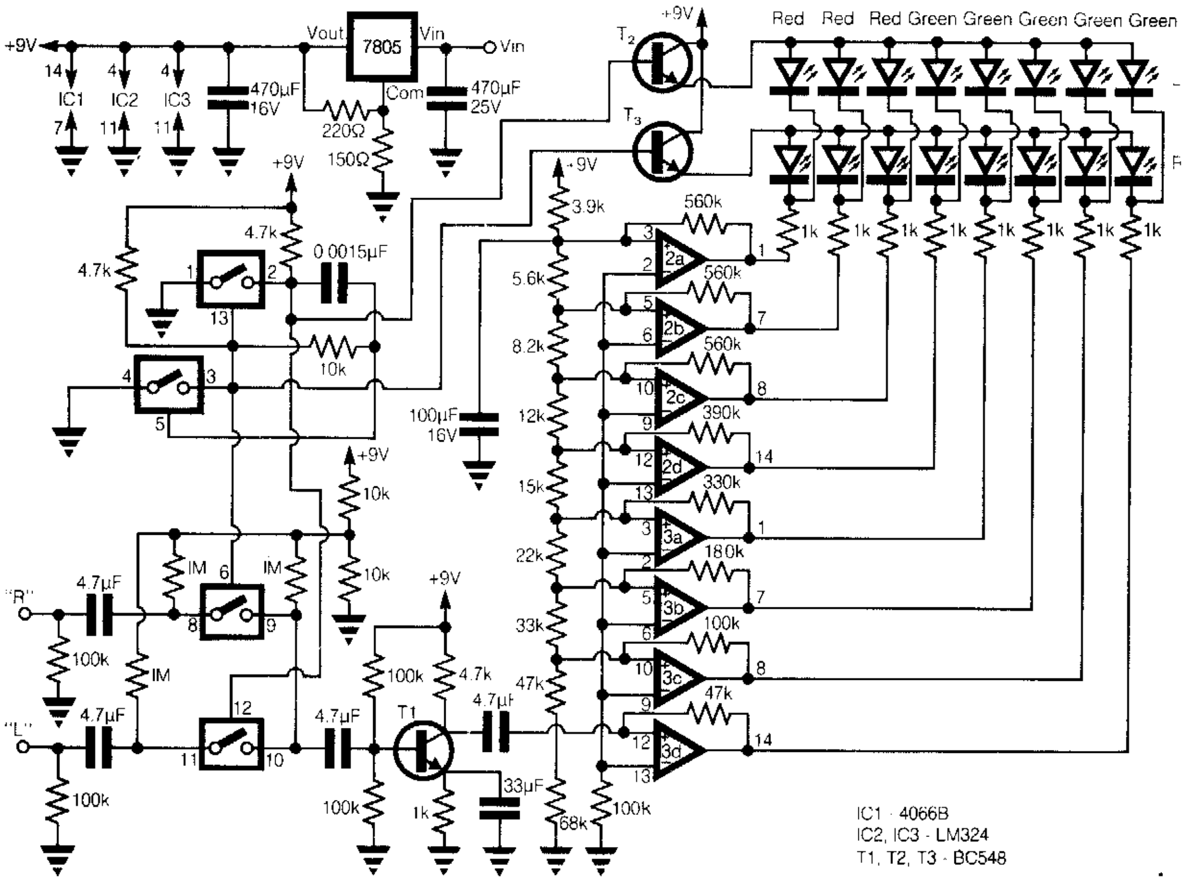
No electrical contact exists between the circuit and the conductor. The 7474 flip-flop is triggered by the output from the Hall sensor, op amp, and Schmitt trigger. This triggering activates the optocoupler, turns on the triac, and trips the circuit breaker.

37

Indicators

The sources of the following circuits are contained in the Sources section, which begins on page 667. The figure number in the box of each circuit correlates to the entry in the Sources section.

STEREO LED VU METER



ELECTRONIC ENGINEERING

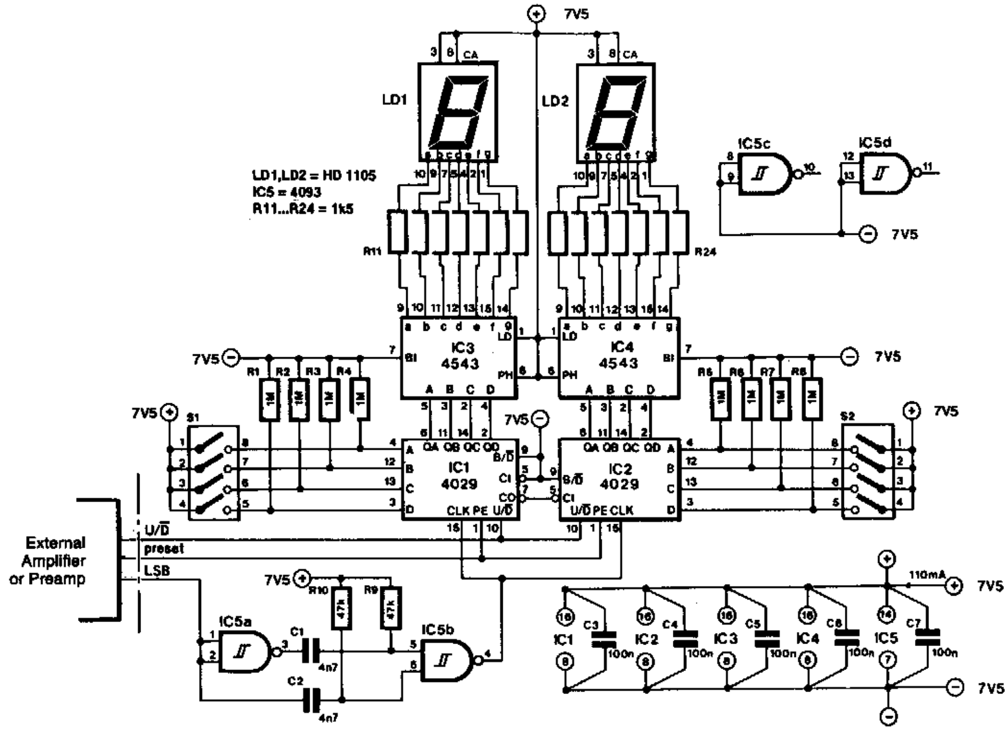
Fig. 37-1

This circuit provides a cheap alternative to the LM3915 series LED displays. The meter relies on a square-wave oscillator built around two CMOS analog switches, which alternatively selects the right and left channels for monitoring and display. The selected signal is amplified by the common-emitter stage T1, and the output is fed into the string comparators which control the display.

These eight comparators are from two LM324 quad op amps, each is connected to a resistor network, which has a 3dB step between each comparator. Each comparator has a positive feedback resistor to increase the hysteresis to provide a longer display, which is switched alternatively at about 10 kHz.

The two CMOS switches in line are biased at half the supply voltage by 1-MΩ resistors from a 100-kΩ divider, which allows them to handle analog signals up to 9 V peak to peak. As the voltage increases above the setpoint of each comparator, the output goes low and the corresponding LED lights, which produces a bar of light in response to the input voltage. For a linear response the resistor-network can be replaced by nine 10-kΩ resistors, giving an equal voltage gap before each LED comes on.

AUDIO AMPLIFIER VOLUME INDICATOR



ELEKTOR ELECTRONICS *Fig. 37-2*

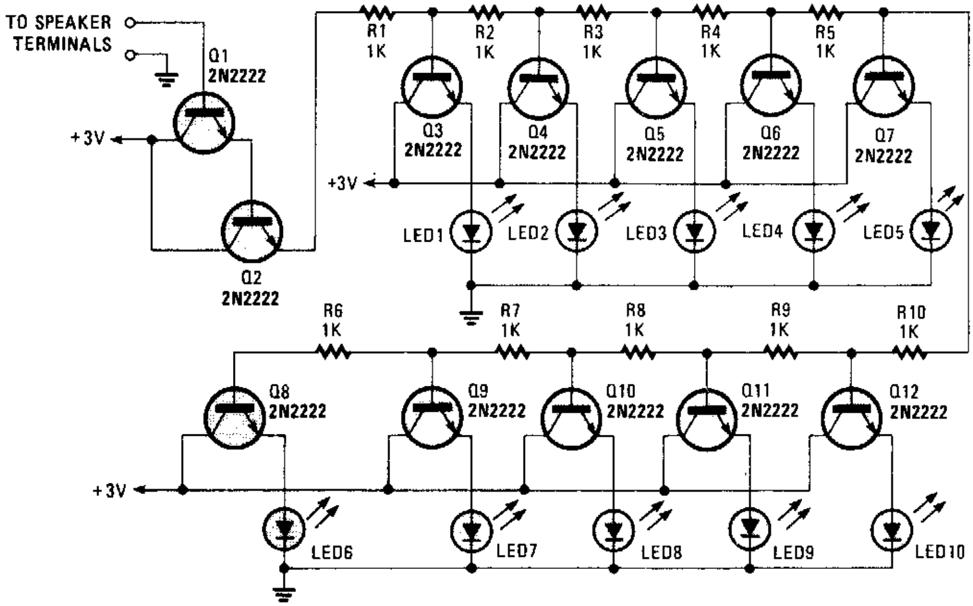
The indicator is intended for use with an audio amplifier or preamplifier, but it can also be used in other applications where a number of steps or changes must be counted rapidly. To prevent interference with the audio signal, the circuit is a static design. Thus, if the volume control is not adjusted, the circuit does nothing.

The circuit does not need an external clock signal, because this is derived from any changes in the least significant bit (LSB). This is done by two differentiating networks: R9/C1 and R10/C2, which double the frequency of an available LSB signal.

Moreover, to ensure that the counters of the indicator remain in step with the volume control, signals "up/down" and "preset" from the preamplifier are used. It might seem rather extravagant to couple the state of the counters in the preamplifier with that of the present counters, but it is a good way to keep the connections between the two units to a minimum. Furthermore, the present counters operate in 8-bit BCD, instead of 6-bit binary as used by those in the volume control (in the preamplifier). All that is required to display the state of the volume control are a couple of BCD-to-7-segment decoders and 7-segment displays.

The preset in the indicator must be set in BCD code. Leading zeros are not suppressed so numbers up to and including 9 are displayed, starting with a 0. The DIP switches and resistors R1 through R8 in the diagram can be omitted if only one fixed preset is likely to be used. The resistors should be replaced by jump leads.

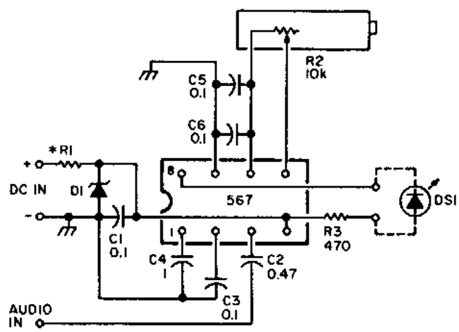
TRANSISTORIZED BAR-GRAPH DRIVER



POPULAR ELECTRONICS *Fig. 37-3*

A resistor network (R1 through R10) with emitter followers (Q1 and Q2) drives LED drivers (Q3 through Q7). This circuit was used as a "light organ" to provide visual volume indication. It can be hooked to a speaker, to another audio source, etc.

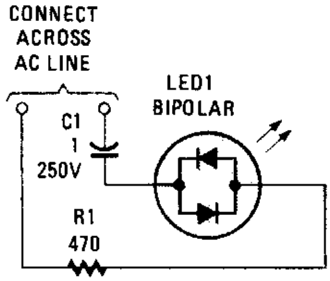
VISUAL CW OFFSET INDICATOR



73 AMATEUR RADIO *Fig. 37-4*

An NE567 tone decoder, tuned to the transceiver's CW offset frequency, ensures that the transceiver will be transmitting on the same frequency as the received CW signal. Simply tune the transceiver so that the LED lights. Eight to 13 Vdc is required; this can be taken from the transceiver supply or an extra battery. Audio is taken from the speaker or headphone output.

ac CIRCUIT LED POWER INDICATOR



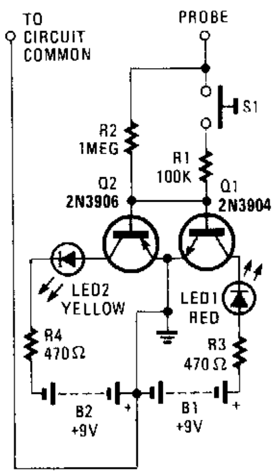
POPULAR ELECTRONICS

Fig. 37-5

Many electronic circuits need an indication that they are under power; for most ac circuits, a neon lamp is the device of choice. A bidirectional tricolor LED can be used if a capacitor is connected in series with the LED to limit the current through the LED. A 1-μF, 250-WVdc capacitor, which has a reactance of 2 650 ohms at 60 Hz, is used in series with an LED to limit the current through the unit to 43 mA. The impedance of the LED is low compared to the reactance of the capacitor, so nearly all the impedance will be caused by the capacitor with the added advantage of no energy loss caused by the capacitor.

The power of the LED is $1.175\ V \times 0.043\ A = 50$ mW compared to an NE-2H at 250 mW. For 230 V, use a 0.47-μF, 400-WVdc capacitor.

ac/dc INDICATOR

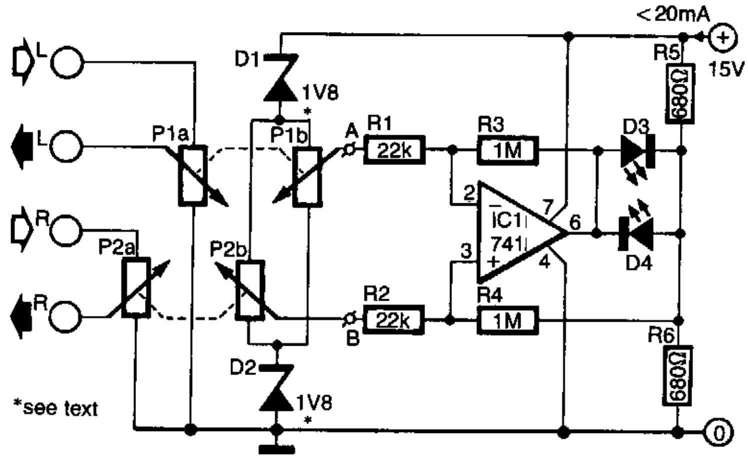


RADIO-ELECTRONICS

Fig. 37-6

By using two switching transistors and two LEDs, this circuit can distinguish low-level ac and dc signals. If the red LED illuminates, the signal is positive dc. If the yellow LED lights, the signal is negative dc. If the signal is ac, both LEDs will light.

BALANCE INDICATOR



ELEKTOR ELECTRONICS *Fig. 37-7*

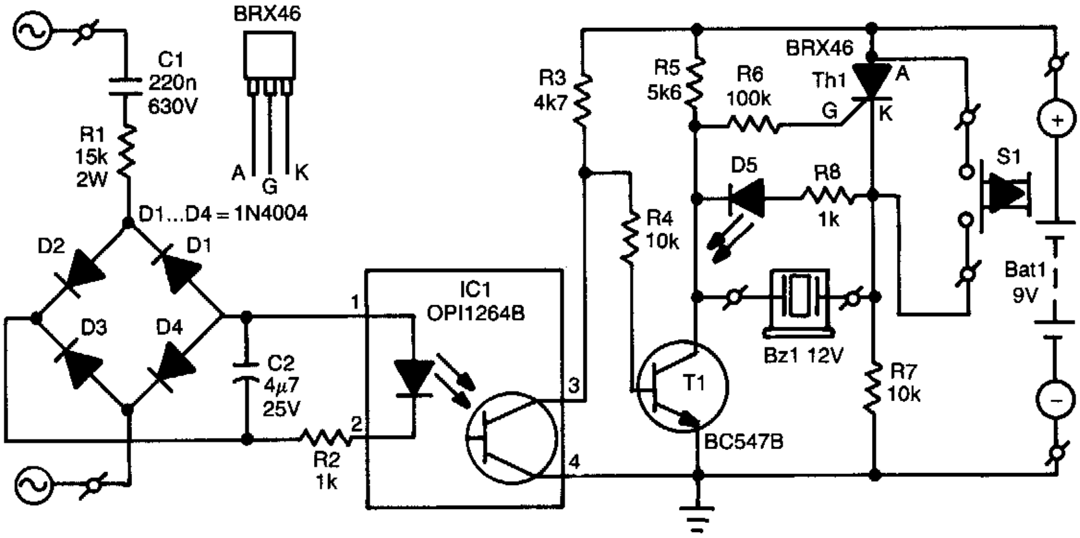
If your amplifier is fitted with two level controls, it actually offers you a balance control and a level control. A drawback of this is that it is quite difficult to set the balance properly. This can be obviated, however, by replacing the two monopotentiometers with stereo versions P1 and P2 in the diagram.

One half of the pair, P1A and P2A, assumes the tasks of the removed components. The other half is connected in a bridge circuit. The voltage between wipers of the potentiometers is then a measure of the balance between the two channels. The lower the potential, the better the balance. If you are interested in knowing the degree of unbalance, connect a center-zero moving coil meter with a bias resistor between A and B. With this arrangement, zener diodes D1 and D2 can be omitted: they are necessary only with the LED indicator shown in the diagram to prevent the input voltage of the op amp from getting too close to the level of the supply voltage.

The circuit around IC1 is a classical differential amplifier. Resistors R5 and R6 provide a virtual earth for the LEDs, which is necessary to ensure that, in spite of the asymmetrical supply voltage, a positive and a negative output is obtained.

Because the LEDs have been included in the feedback loop of the indicator, the circuit is pretty sensitive. At only 40 mV, that is, just 1/400 of the supply voltage, one of the LEDs begins to light. The maximum current drawn by the LEDs is determined by the values of R5 and R6.

MAINS FAILURE INDICATOR



ELEKTOR ELECTRONICS

Fig. 37-8

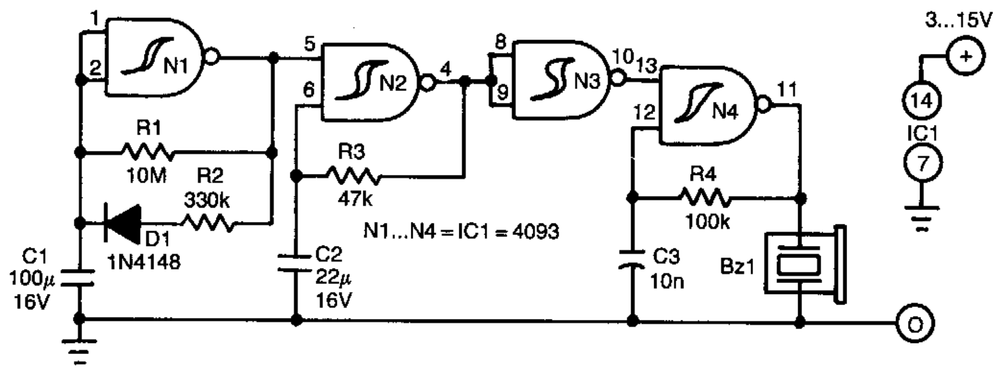
When the mains voltage is present at the input terminals, the transistor in the optocoupler is on, T1 is off, and silicon-controlled rectifier Th1 is in the conducting state. Because both terminals of the piezoelectric buzzer are then at the same potential, the buzzer is inactive. If the mains voltage drops out, transistor T1 conducts and causes one of the terminals of the buzzer to be connected to earth; the thyristor remains in the conducting state. In this situation, a large enough potential difference is across both the buzzer and D5 to cause these elements to indicate the mains failure—both audibly and visibly.

When the mains is restored, the circuit returns to its original state. A touch on the reset button then interrupts the current through the SCR so that the thyristor goes into the blocking state, and the other terminal of the buzzer is connected to ground.

The unit is powered by a 9-V battery and draws a quiescent current of 1.7-2.5 mA. It is important for the enclosure to be well-insulated.

If by accident the circuit to the optocoupler and R2 is broken, electrolytic capacitor C2 might be damaged because it will be charged well above its 25-V rating. Secondly, where a plug is used for the mains connection, it is advisable to solder a 1-MΩ resistor across C1 so that this capacitor does not retain its charge after the plug is removed from the mains socket.

ON INDICATOR



ELEKTOR ELECTRONICS

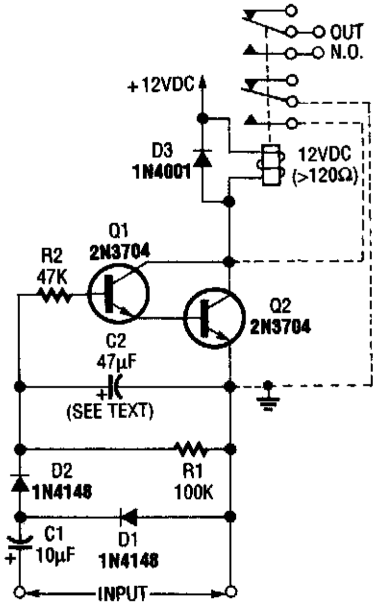
Fig. 37-9

Battery-operated equipment can work on one set of batteries for a long time nowadays. However, if it is left on inadvertently, that "long time" is over very quickly. Moreover, flat (dead) batteries are always found at the wrong moment. The circuit proposed here is a sort of ***aide-memoire***. Every two minutes, it emits 5 to 10 pips to indicate that the equipment is still switched on.

Basically, the circuit consists of three rectangle-wave generators and an inverter. The first of the generators is formed by N1 and provides a signal with a period of about two minutes and a pulse duration of around 10 seconds. During those 10 seconds, the second generator starts operating in a one-second rhythm. Thus, N2 outputs 10 pulses every 2 minutes. That output is inverted so that N4, like N2, can only be enabled during the 10-second pulse train from N1. The difference is that during those 10 seconds, N4 is enabled and inhibited 10 times; this is what causes the pips.

Do not take the times and number of pulses too literally, because wide variances are between ICs from different manufacturers. On the other hand, component values are not critical, so it is fairly easy to adapt the circuit to personal taste or requirements. The buzzer can be a standard Toko type or equivalent. The current drawn by the circuit is negligible.

SOUND SENSOR



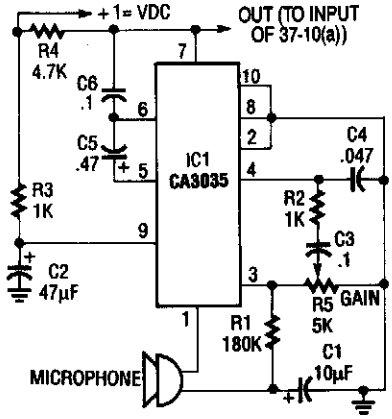
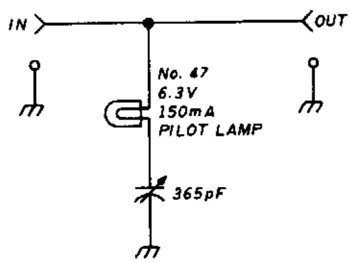


Fig. 37-10(b)

RADIO-ELECTRONICS *Fig. 37-10(a)*

By using a microphone, high-gain amplifier (Fig. 37-10(b)), and detector-relay driver (Fig. 37-10(a)) a sound-detecting alarm system can be constructed. If you want a latching setup, make the dotted connections to the relay shown in Fig. 37-10(a).

TRANSMITTER OUTPUT INDICATOR



HAM RADIO *Fig. 37-11*

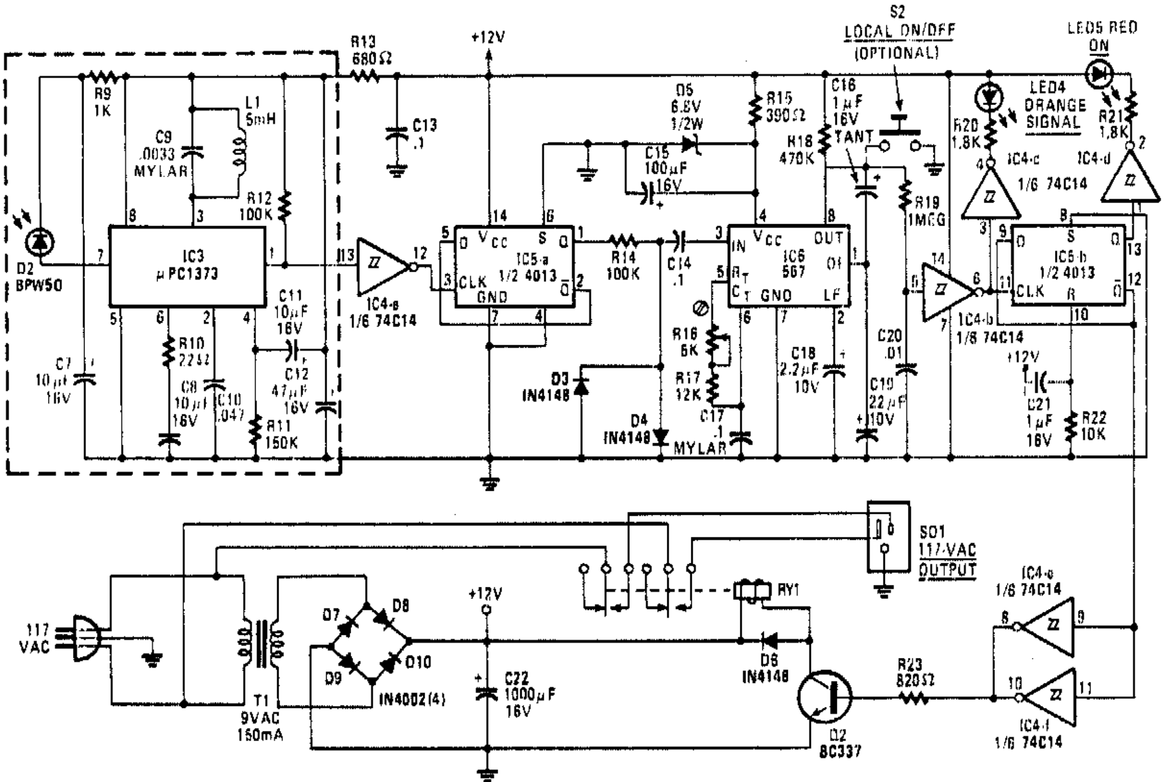
Relative power can be indicated with this simple circuit. Adjust the 365-pF variable capacitor for desired lamp brightness.

38

Infrared Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 667. The figure number in the box of each circuit correlates to the entry in the Sources section.

IR RECEIVER I



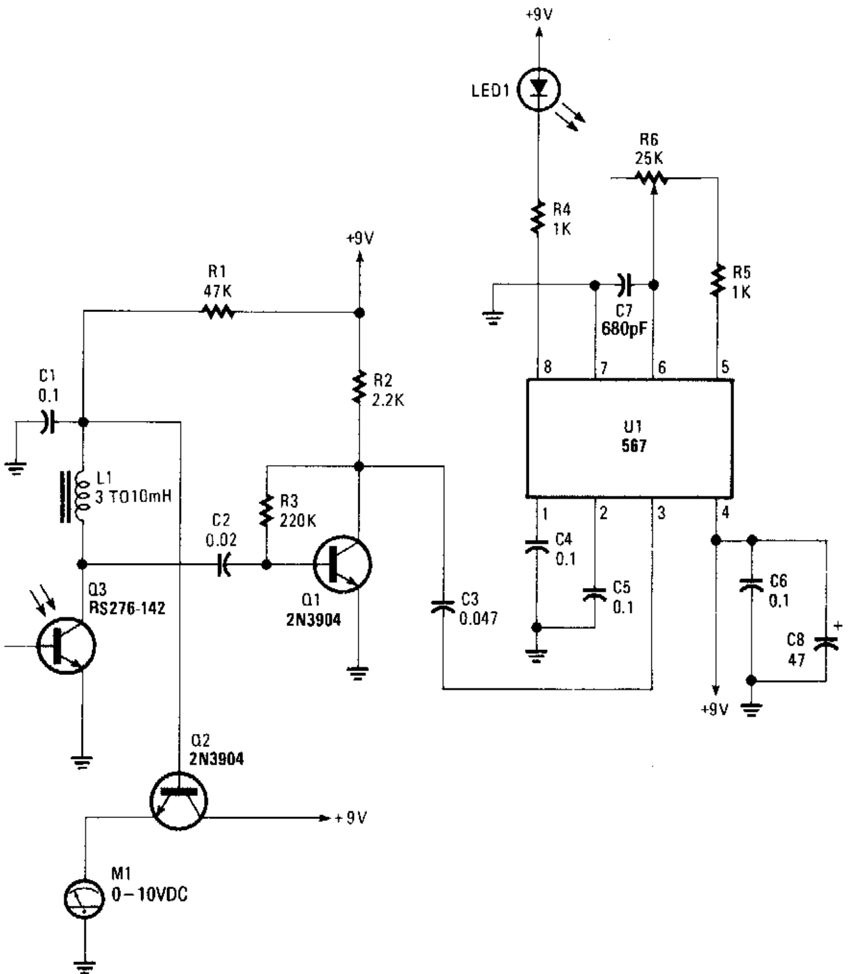
RADIO-ELECTRONICS

Fig. 38-1

This receiver is built around a uPC1373 IR remote-control preamplifier, a sensitive 30-to-40 kHz tuned detector, an automatic gain control, a peak detector, and an output waveshaping buffer. The demodulated signal from the preamp stage is sent to IC4A, a 74C14 Schmitt trigger. The squared-up 1 500-Hz signal is then sent to the clock input of IC5A, half of a 4013 dual "D" flip-flop. That 750-Hz signal is clipped to approximately 0.7-V p-p by diodes D3 and D4. The clipped signal is then fed to IC6, a 567 tone decoder. The output of that IC goes low whenever the frequency of the signal fed to it is within the lock range of its internal VCO.

When IC6 detects a signal of the proper frequency, pin 8 goes low. The output signal is fed through another Schmitt trigger (IC4B), which drives another "D" flip-flop, IC5B. Schmitt trigger IC4B also drives IC4C, which in turn drives LED4, SIGNAL, which lights up whenever a signal is received. The Q output of IC5B drives two parallel-connected inverters. IC4C and IC4F turn transistor Q2 on when Q goes low. That transistor energizes the relay; its contacts switch the controlled device on and off.

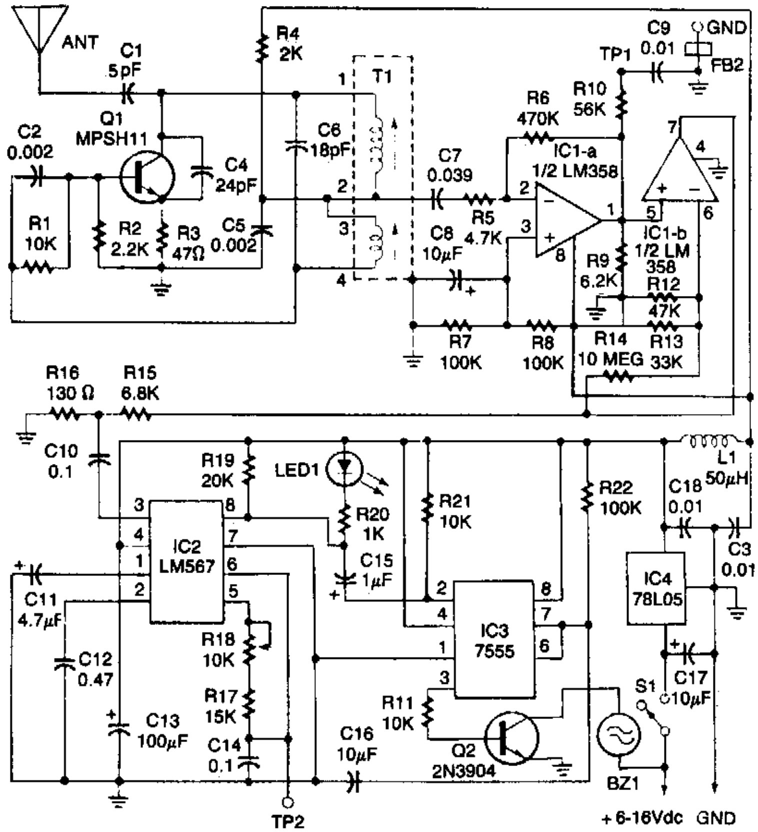
IR RECEIVER II



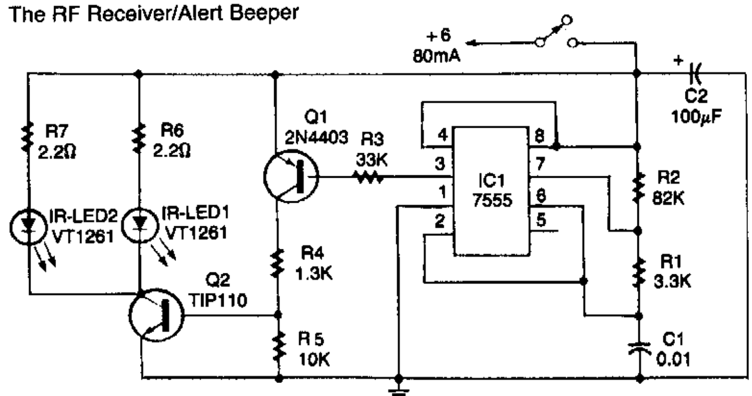
POPULAR ELECTRONICS ***Fig. 38-2***

Q3 is an IR phototransistor that responds to a modulated IR beam. Q1 amplifies the ac component of the IR beam. Q2 drives a meter as a relative indication of the strength of the light beam. A strong beam gives a lower meter reading. U1 is a tone decoder that produces a low output on pin 1 during reception for an IR beam that is modulated with the correct tone frequency, determined by R6.

WIRELESS IR SECURITY SYSTEM



The RF Receiver/Alert Beeper

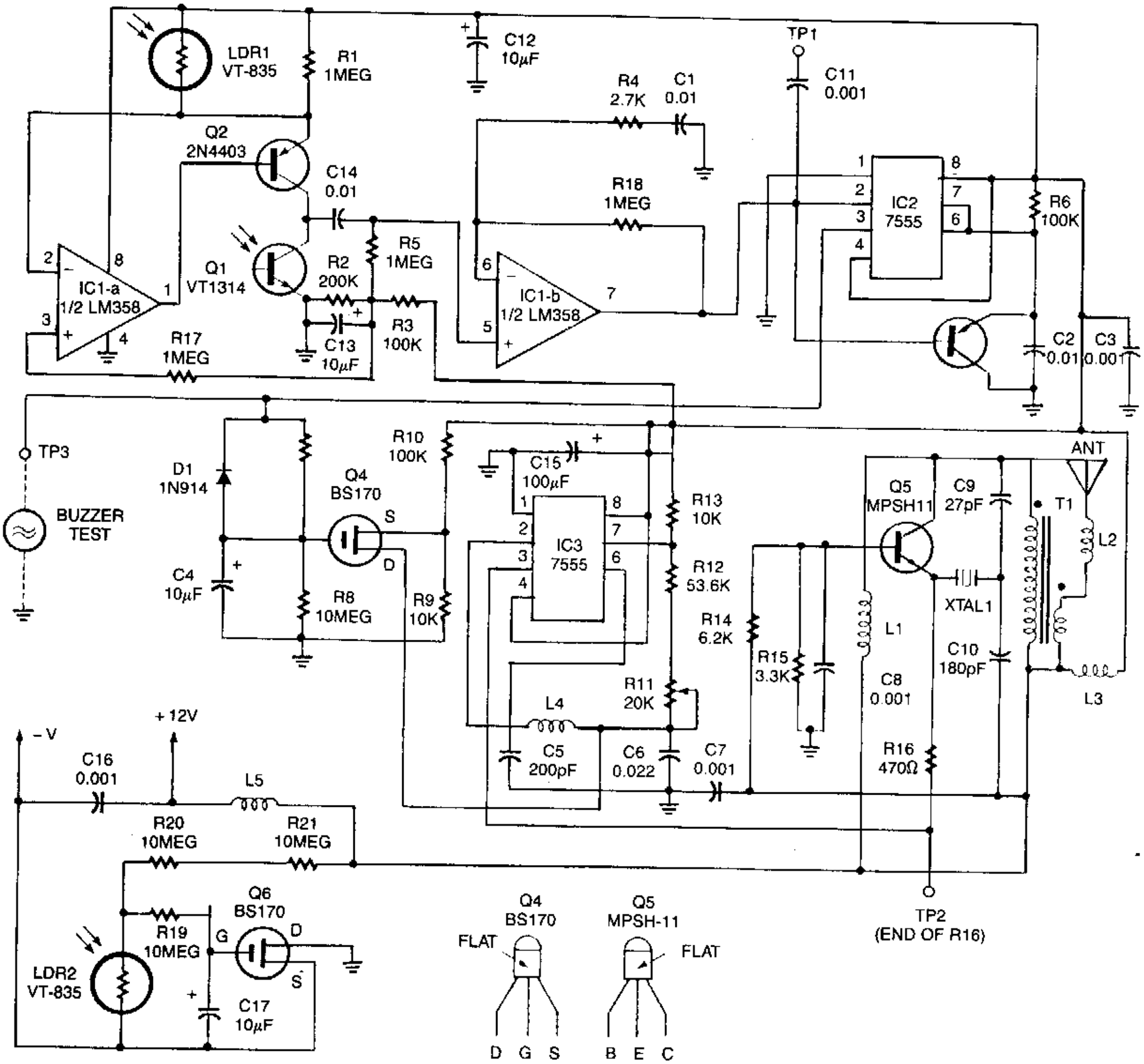


RADIO-ELECTRONICS

Fig. 38-3

The IR Transmitter

This system contains an IR transmitter, an IR receiver/RF transmitter, and an RF receiver/alert beeper. Two IR LEDs in the transmitter transmit a pulsed beam of invisible infrared light to the receiver, which contains an IR phototransistor. The phototransistor detects and amplifies the pulse-modulated IR beam. If the receiver section senses that the IR beam is momentarily interrupted by an object blocking the

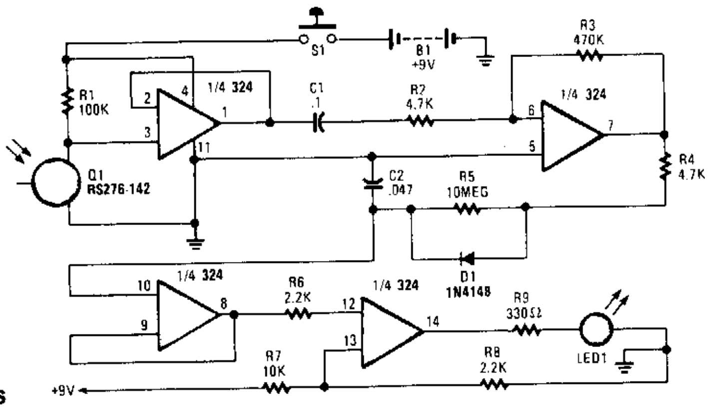


The IR Receiver/RF Transmitter

beam's path, it triggers the transmitter, which outputs a 49.890-MHz carrier that is amplitude-modulated by a 490-Hz tone.

Upon receiving the 490-Hz amplitude-modulated carrier, the RF receiver/beeper unit sounds an alarm that alerts the user to the intrusion. The system is not limited to just one RF transmitter. A single RF receiver/beeper can be used to monitor any number of RF transmitters (or locations). However, the receiver/beeper unit cannot discriminate between different transmitter sites in multiple-transmitter systems.

IR DETECTOR

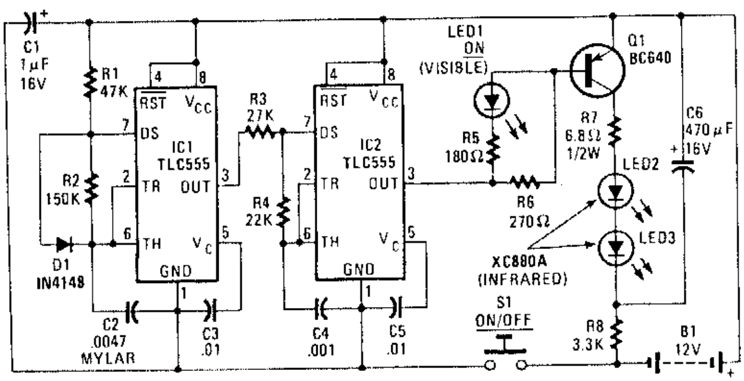


POPULAR ELECTRONICS

Fig. 38-4

Useful for checking TV remote controls, IR-based alarm systems, and IR sources, this circuit causes LED1 to turn on for two seconds in the presence of IR light pulses. U1A acts as a voltage follower for detector Q1. C1 and R2 form a differentiating network and U1B acts as an amplifier for the pulses, which charges C2. Voltage follower U1C samples the voltage on C2 and drives comparator U1D, which switches LED1 on or off.

INFRARED REMOTE CONTROLLER



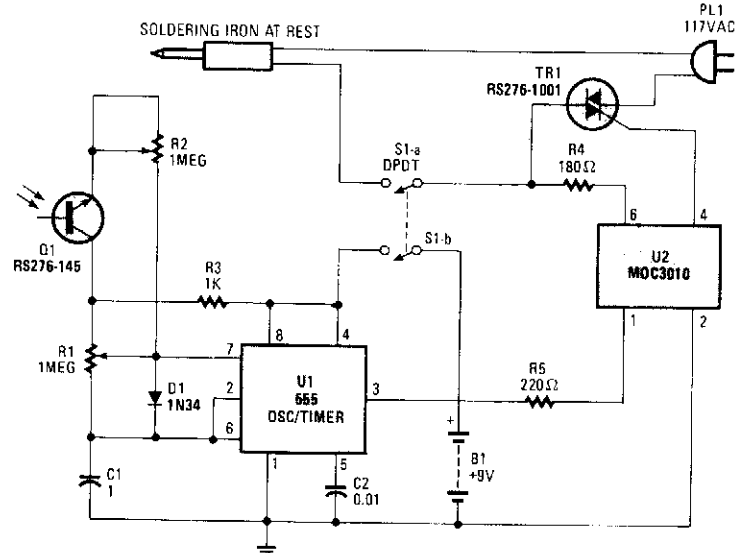
RADIO-ELECTRONICS

Fig. 38-5

The transmitter is built around two CMOS 555 timer ICs (TLC 555s). The transmitter generates a modulated 35-kHz IR signal. The 35-kHz carrier frequency is generated by IC2, and the 1 500-Hz modulating signal is generated by IC1. The output of IC2 drives LED1 through resistor R5; that LED provides visual indication that the transmitter is working. In addition, IC2 drives transistor Q1, which drives the two infrared LEDs (LED 2 and LED3).

To provide the high current needed to drive the two IR LEDs, capacitor C6 is precharged, the charge it contains is dumped when S1 is pressed. When S1 is not pressed, the power to the ICs is cut off. However, C6 is kept charged via R8. Then, when S1 is pressed, the current stored in C6 can be used to drive the LEDs for as much as 1/2 second. That's plenty of time for the receiver to pick up a signal.

IR-CONTROLLED SOLDERING STATION



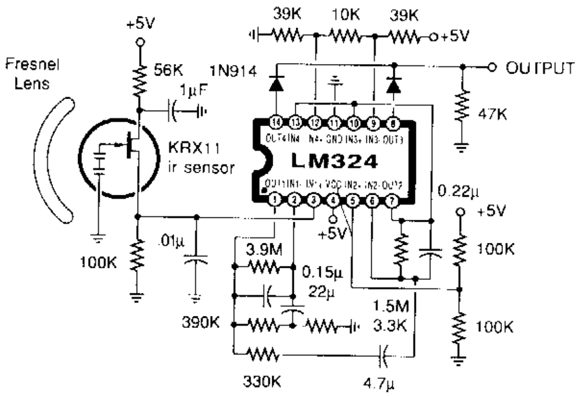
POPULAR ELECTRONICS

Fig. 38-6

An IR-sensitive phototransistor is used to sense soldering-iron temperature. The phototransistor must see the tip through an opaque tube to avoid stray light, and the phototransistor should preferably be fitted with an IR filter. An old photo negative, dark red plastic, or red or black glass can be used. The iron sits on a holder.

When the iron is removed from the holder, the iron is not being viewed by the detector. The heat will increase, but the circuit has a lag time; if the iron is returned to its holder after each use, overheating should not be a problem.

INFRARED "PEOPLE" DETECTOR

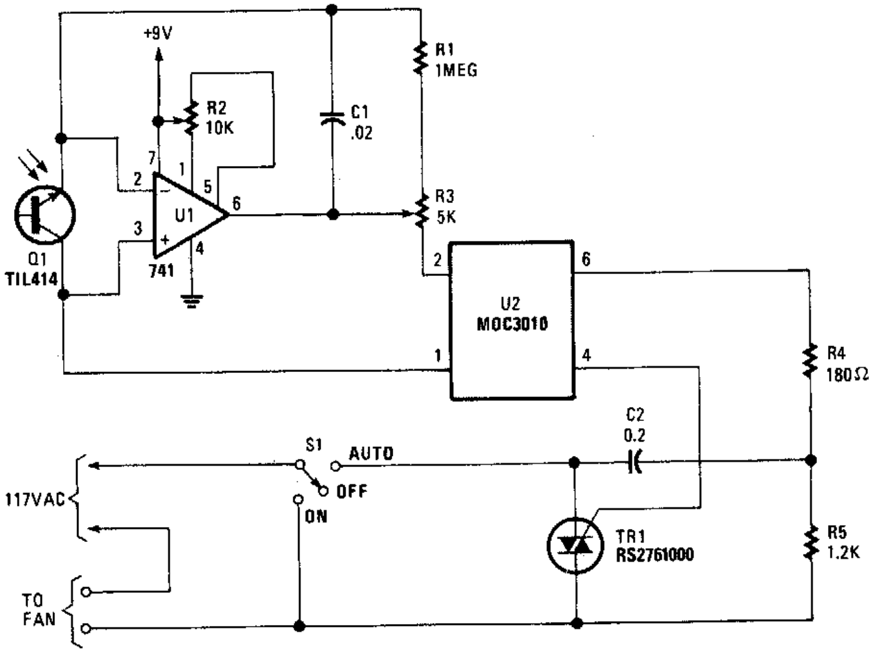


This circuit uses an Amperex pyroelectric IR sensor, an LM324 op amp (configured as a high-gain amplifier in the 0.3-to 5-Hz range), and a window detector. The output will go high on any motion, which will change the infrared signature seen by the sensor.

RADIO-ELECTRONICS

Fig. 38-7

IR HEAT-CONTROLLED KITCHEN FAN

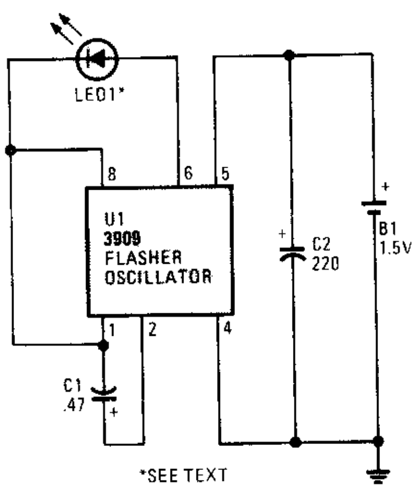


POPULAR ELECTRONICS

Fig. 38-8

Q1 senses IR from heat sources, causes U1 to switch, activates optocopuler U1, and triggers TR1. This controls a fan. The Triac is from Radio Shack, or else a 200-V, 6-A unit (C106B) can be used.

SIMPLE IR TRANSMITTER



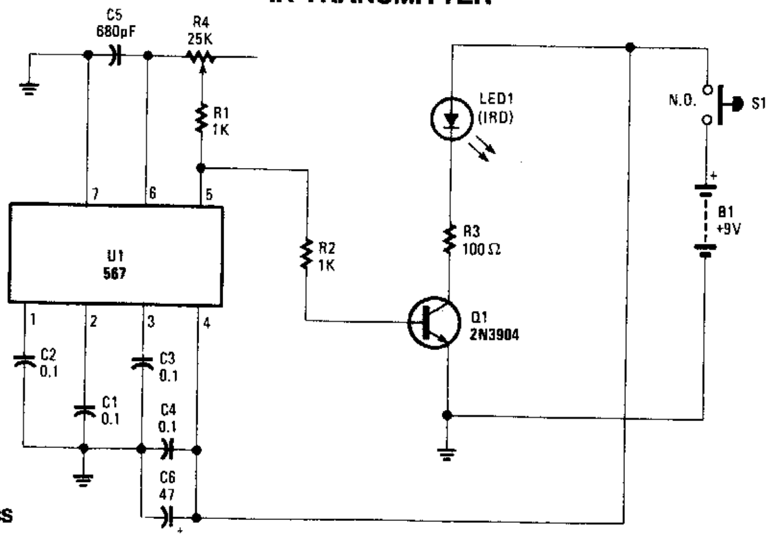
POPULAR ELECTRONICS

Fig. 38-9

The IR diode's flash rate is determined by the value of C1, a 220-μF capacitor that sets the rate of oscillation at 1 Hz per second. Reducing C_1 will increase the frequency of the circuit, while larger values will decrease the frequency.

Because the circuit only sends out single, narrow pulses of invisible light, the IR receiver only responds with a click for every output pulse.

IR TRANSMITTER

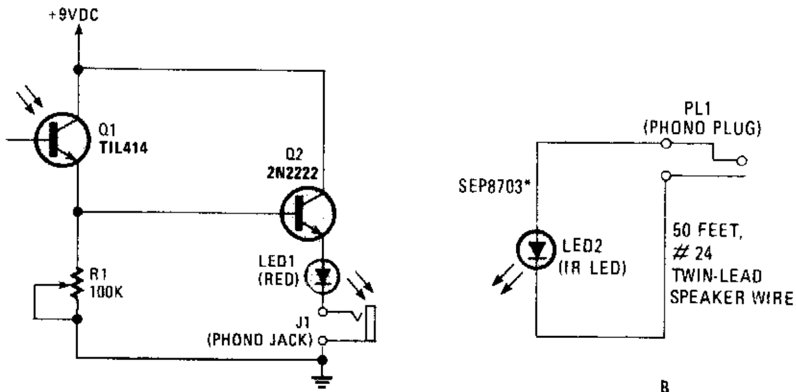


POPULAR ELECTRONICS

Fig. 38-10

Using an NE567 as a tone oscillator, this circuit produces an IR signal from the LED, which is modulated with a square wave. LED1 is an IR-emitting LED. The modulation helps improve performance under high ambient light conditions.

IR REMOTE EXTENDER



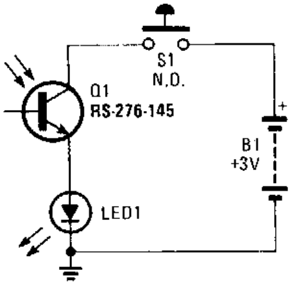
POPULAR ELECTRONICS

Fig. 38-11

This circuit can be used to operate a VCR or CD player from another room. It's really an infrared signal repeater. The signal from the remote is received and then retransmitted over wires to an infrared LED. The beam from the LED is then picked up by the receiving window on the VCR or CD player.

The visible light LED (LED1) in series with the IR unit (LED2) is used to indicate that the transmitted signal has been detected. The 100-kΩ trimmer potentiometer (R1) adjusts the repeater's sensitivity. The resistor that is usually found in series with the LEDs is omitted, because the voltage reading is about 1.0 Vdc as a result of the voltage drop across the lines.

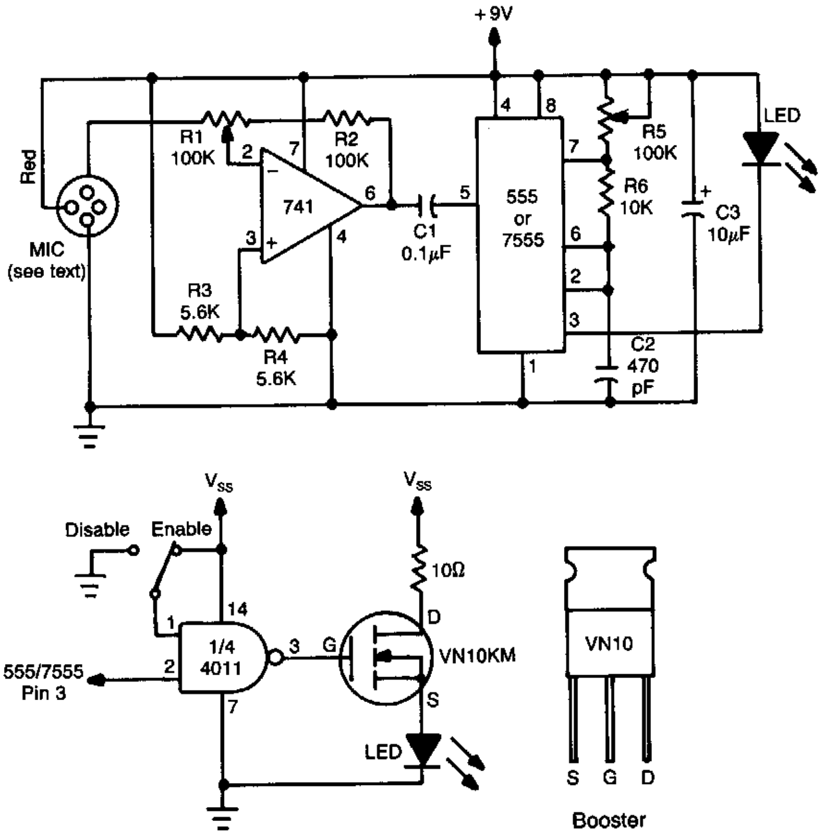
INFRARED REMOTE CONTROL TESTER



Using a battery, a phototransistor and a visible-light LED, this simple circuit is a ''go/no go'' tester for IR remote control devices. The illumination of the LED indicates that Q1 is being modulated by IR energy.

POPULAR ELECTRONICS *Fig. 38-12*

VOICE-MODULATED PULSE FM IR TRANSMITTER



POPULAR ELECTRONICS *Fig. 38-13*

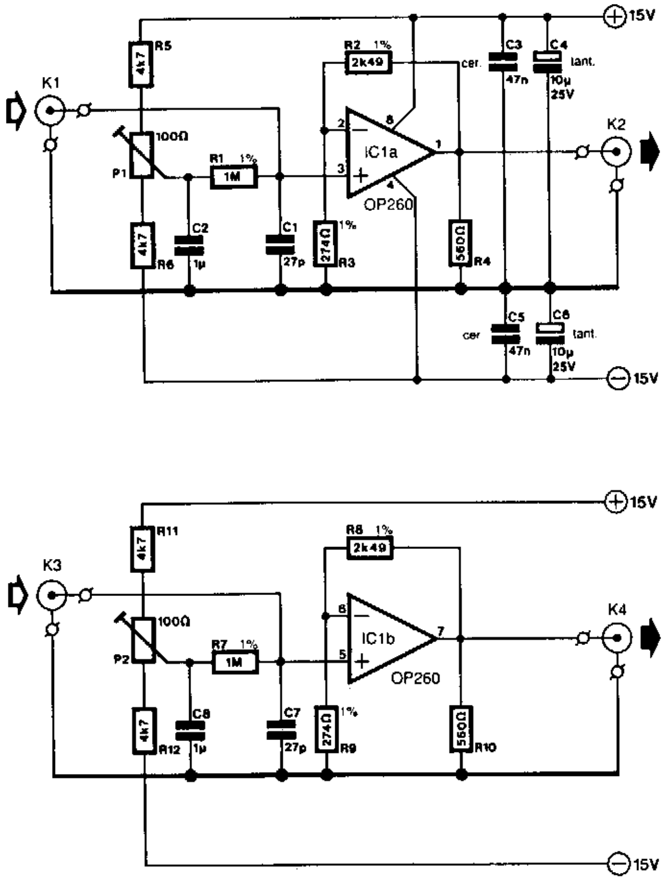
This circuit has a 741 audio amplifier, which is fed by a microphone (use an amplified type), an FM modulator, and a CMOS timer that acts as a VCO. The LED is pulsed with the timer output (the booster circuit can be used for increased range). This yields an FM-modulated, pulsed IR beam.

39

Instrumentation Amplifiers

The sources of the following circuits are contained in the Sources section, which begins on page 668. The figure number in the box of each circuit correlates to the entry in the Sources section.

OSCILLOSCOPE PREAMP



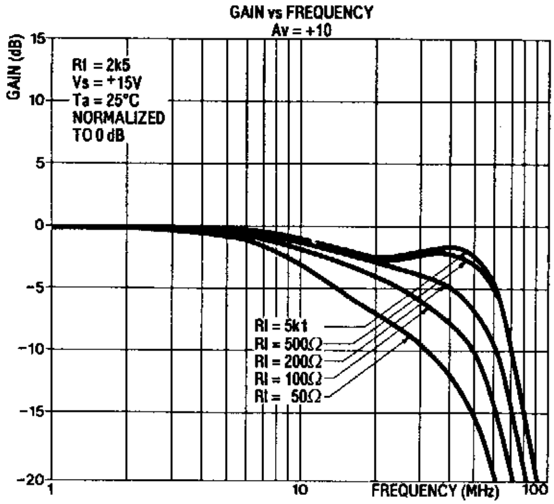
ELEKTOR ELECTRONICS

Fig. 39-1(a)

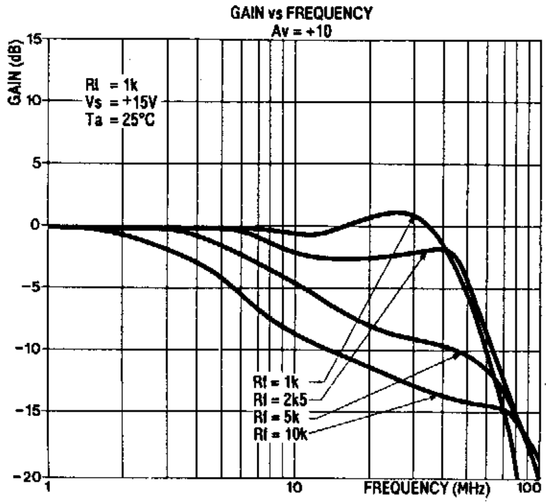
In many oscilloscopes, the most sensitive range is 2 to 5 mV, although it is often possible to improve this to 1 to 2 mV by a variable gain control. To obtain even better sensitivity, the present preamplifier, which has an amplification of about 10 (20 dB), might be useful.

Because most oscilloscopes have a bandwidth of 20 MHz or more, the amplifier must, of course, have a slightly wider bandwidth and that is achieved with a Type OP260 op amp. This has a slew rate of 550 V/μs (at an amplification of 10) and a bandwidth of 40 MHz that is virtually independent of the amplification. The gain vs. frequency response is not so good, however: as can be seen from Fig. 39-1(b), where the characteristics are given for a number of loads. The hump in the curves depends on the value of the feedback resistor, whose optimum value appears to be 2.5 kΩ.

The curves in Fig. 39-1(c) accord with different values of R_2/R_8 for an amplification factor of 10. Some experimentation with the value of R_2/R_8 for different amplification factors can be instructive. Remember, however, that the output impedance increases from 20 to 225 Ω over the frequency range of 10 MHz to 60 – 70 MHz. It is therefore important to keep all connections on the prototyping board as short as possi-



ELEKTOR ELECTRONICS *Fig. 39-1(b)*



ELEKTOR ELECTRONICS *Fig. 39-1(c)*

ble and to connect all earth points to a common ground via a separate, heavy track. Also, do not use an IC socket.

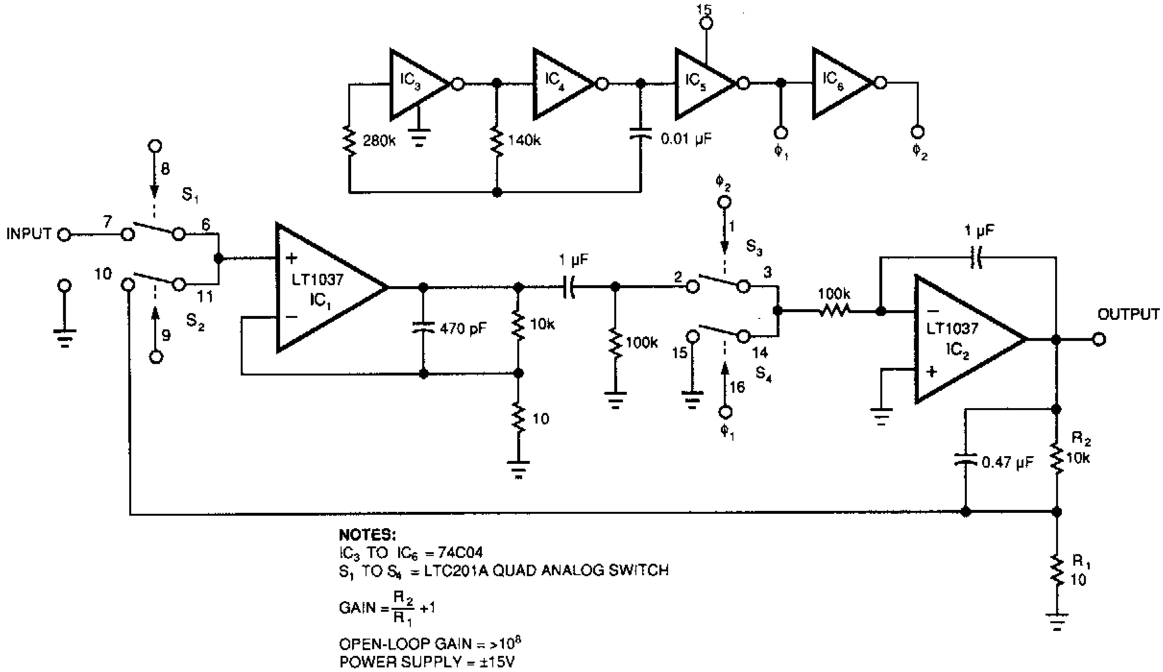
An input impedance of 1 MΩ was chosen, which results in a fairly high level of noise at the output (with open-circuit input). This value can be reduced, because otherwise the use of a 1:10 probe will be inhibited; it would give constant problems with the noise. However, when the amplifier is connected to a suitable source, the noise reduction is normally more than ample to obtain a good trace on the screen. Presets P1 and P2 provide compensation for the dc offset and input offset, caused by R1 and R7 respectively.

The input bias current for the noninverting input is about 10 times lower than that for the inverting input, which makes the OP260 more suitable for noninverting circuits. The inverting circuit can also give problems because of the low values of R_2 (R_8) and R_3 (R_9). The input bias current is typically 0.2 μA, and the input offset is about 3 mV (max. 7 mV).

In this type of circuit, it is important to use a well-regulated power supply. The power-supply suppression up to 10 kHz is roughly 70 dB, and this reduces with increasing frequency. Any noise or tiny ripple on the supply lines would make the application of the circuit as a small-signal amplifier impossible.

The circuit draws a current of about 14 mA. The slew rate, as with most op amps, is asymmetric and might lead to visible distortion of the signal when the drive to the 560-Ω resistor is high at the higher frequencies.

LOW-DRIFT/LOW-NOISE dc AMPLIFIER



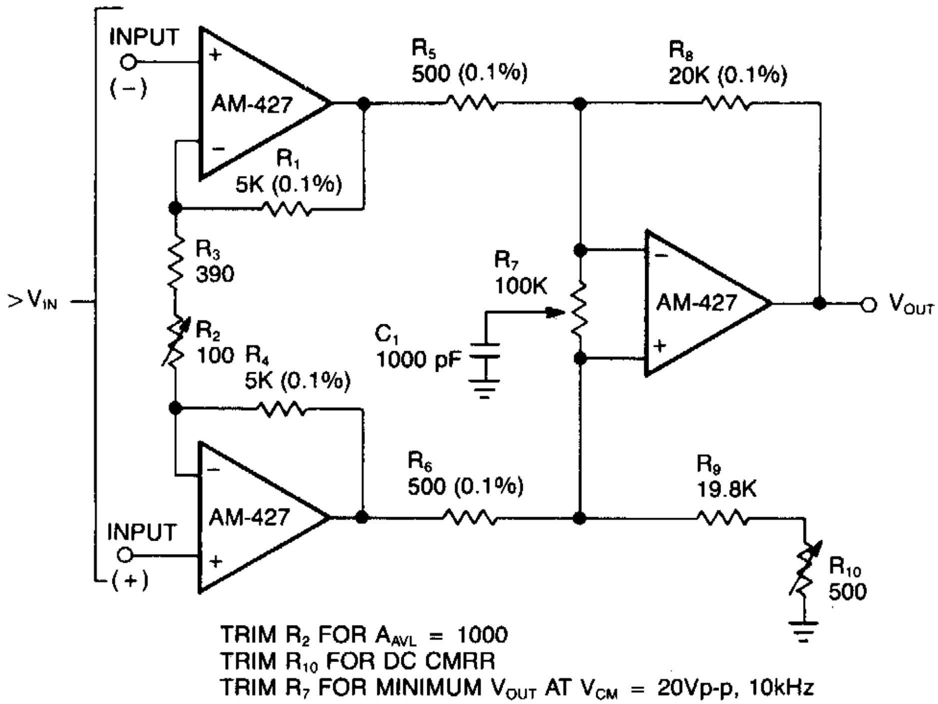
EDN

Fig. 39-2

Figure 39-2's circuit combines a low-noise op amp, IC1, with a chopper-based carrier-modulation scheme to achieve a low-noise, low-drift dc amplifier whose performance exceeds any currently available monolithic amplifier. The amplifier's offset is less than 1 μV, and its drift is less than 0.05 μV/°C. This circuit has noise within a 10-Hz bandwidth less than 40 nV. The amplifier's bias current, which is set by the bipolar input of IC1, is about 25 nA.

The 74C04 inverters (IC3 to IC6) form a simple 2-phase square-wave clock that runs at about 350 Hz. The complementary oscillator signals (O_1 and O_2) provide drive to S1 and S2, respectively, causing a chopped version of the input to appear at IC1's input. IC1 amplifies this ac signal. S3 and S4 synchronously demodulate IC1's square-wave output. Because S3 and S4 switch synchronously with S1 and S2, the circuit presents proper amplitude and polarity information to IC2, the dc output amplifier. This output stage integrates the square wave to provide a dc voltage output. R1 and R2 divide the output and feed it back to the input chopper where the divided output serves as a zero signal reference. The ratio of R1 and R2 sets the gain, in this case to 1 000.

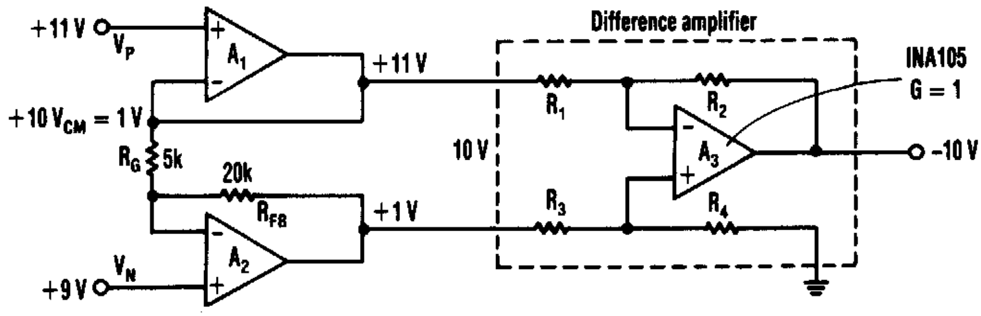
INSTRUMENTATION AMPLIFIER



DATEL *Fig. 39-3*

In a single difference amplifier configuration, the AM-427 exhibits excellent common-mode rejection and spot noise voltage so low it is dominated by the resistor Johnson noise. The three-amplifier configuration shown avoids the low input-impedance characteristics of difference amplifiers. Because of the additional amplifiers used, the spectral noise voltage will increase from a typical of 3 nV/Hz to approximately 4.9 nV/Hz. The overall gain of the circuit is set at 1 000; with balanced source resistors, a CMRR of 100 dB is achieved.

EXTENDED COMMON-MODE INSTRUMENT



ELECTRONIC DESIGN

Fig. 39-4

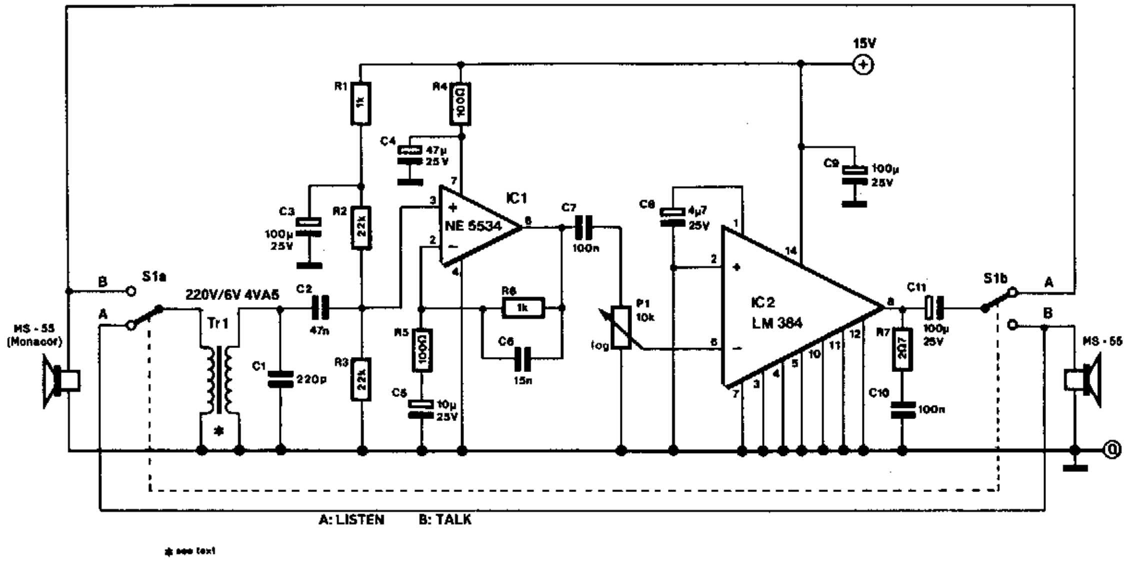
These circuits allow a larger common-mode range than most instrument amplifier inputs can allow.

40

Intercom

The source of the following circuit is contained in the Sources section, which begins on page 668. The figure number in the box correlates to the entry in the Sources section.

Two-Wire Intercom



ELEKTOR ELECTRONICS *Fig. 40-1*

The design consists of an amplifier, a double-pole changeover switch and two loudspeakers: one for the master station and one for the slave. More than one slave unit can be used, but each requires an additional changeover switch.

The power amplifier is a Type LM384, which can provide almost 2 W output at a supply voltage of 15 V. Pins 3, 4, 5, 10, 11, and 12 are connected to ground and at the same time afford some cooling of the device. Because of that, the IC should not be fitted in a socket, but be soldered direct to the circuit board.

The LM384 processes signals with respect to earth so that an asymmetric supply is sufficient. The amplification has been set internally to ×50 (34 dB). The IC's supply line is decoupled by C9.

To ensure adequate input sensitivity, a preamplifier, IC1, is provided, which has an amplification of 11 (21 dB). Because this stage is intended for speech only, its bandwidth is limited to 160 Hz to 10 kHz. Divider R2/R3 at the input of the op amp is decoupled by C3.

Special loudspeakers that can also serve as microphones are readily available: in the prototype, MS-55 units from Monacor were used, but a number of other makes will do just as well. The bandwidth of the MS-55 (used as loudspeaker) extends from 150 Hz to 20 kHz and (used as a microphone) from 20 Hz to 20 kHz. The MS-55 can handle up to 5-W output. To ensure satisfactory operation, particularly as a microphone, the loudspeaker must be fitted in a closed box.

Although it is advantageous for the "microphone" to have a low internal resistance, it is necessary for a transformer to be used at the input of the circuit. This has, however, the advantage that long cables can be used. The present circuit uses a standard mains transformer instead of a special microphone transformer. For this purpose, the secondary (6 V) winding is connected to the "microphone." The microphone impedance is thereby magnified from about 8 Ω to around 10 kΩ. The power handling of the transformer is quite high to ensure that signal losses in the primary winding are kept at a minimum. Capacitor C1 suppresses HF interference.

If the mains transformer and the "microphone transformer" are housed in the same enclosure, some trial and error and screening are necessary to eliminate hum. The "microphone transformer" itself might cause hum in the remainder of the circuit. In that case, the preamp stage must also be screened.

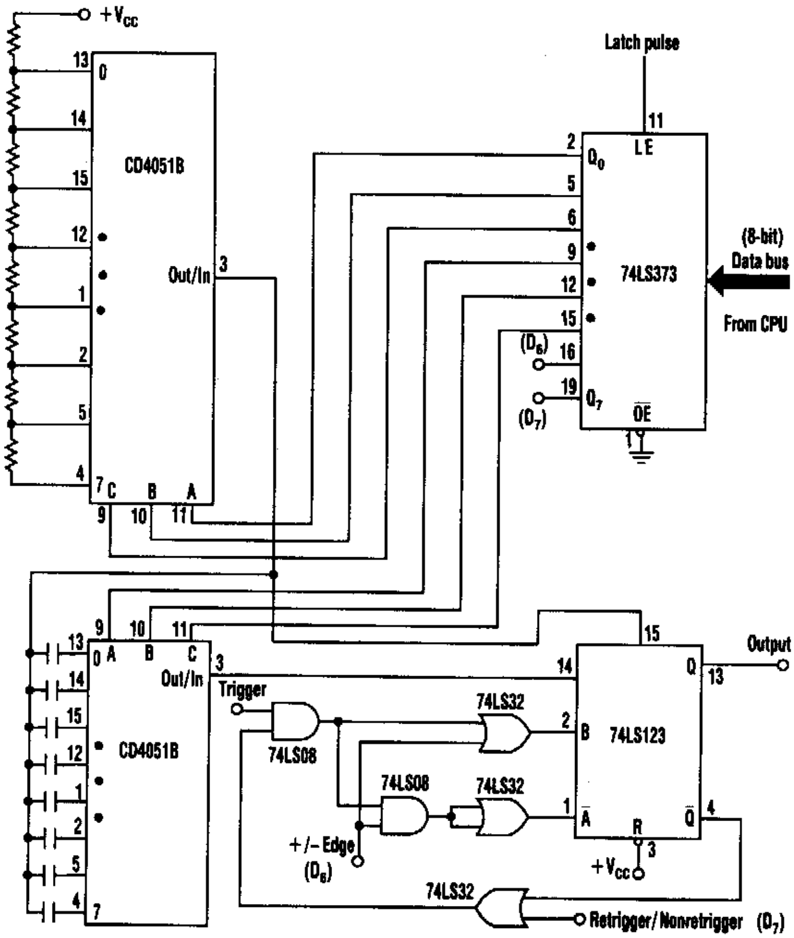
In the prototype, the speech bandwidth was limited from 400 Hz to 4 kHz and this proved perfectly acceptable for good speech transfer. Most of the current drawn by the circuit flows through the power amplifier. At worst, this is 210 mA (680 mA peak), when the amplifier delivers 1.8-W output. The LM384 can deliver a power of up to 5 W. The supply voltage should then be raised to 22 V and a heatsink for the device will be necessary.

41

Interface Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 668. The figure number in the box of each circuit correlates to the entry in the Sources section.

VERSATILE ONE-SHOT CPU INTERFACE



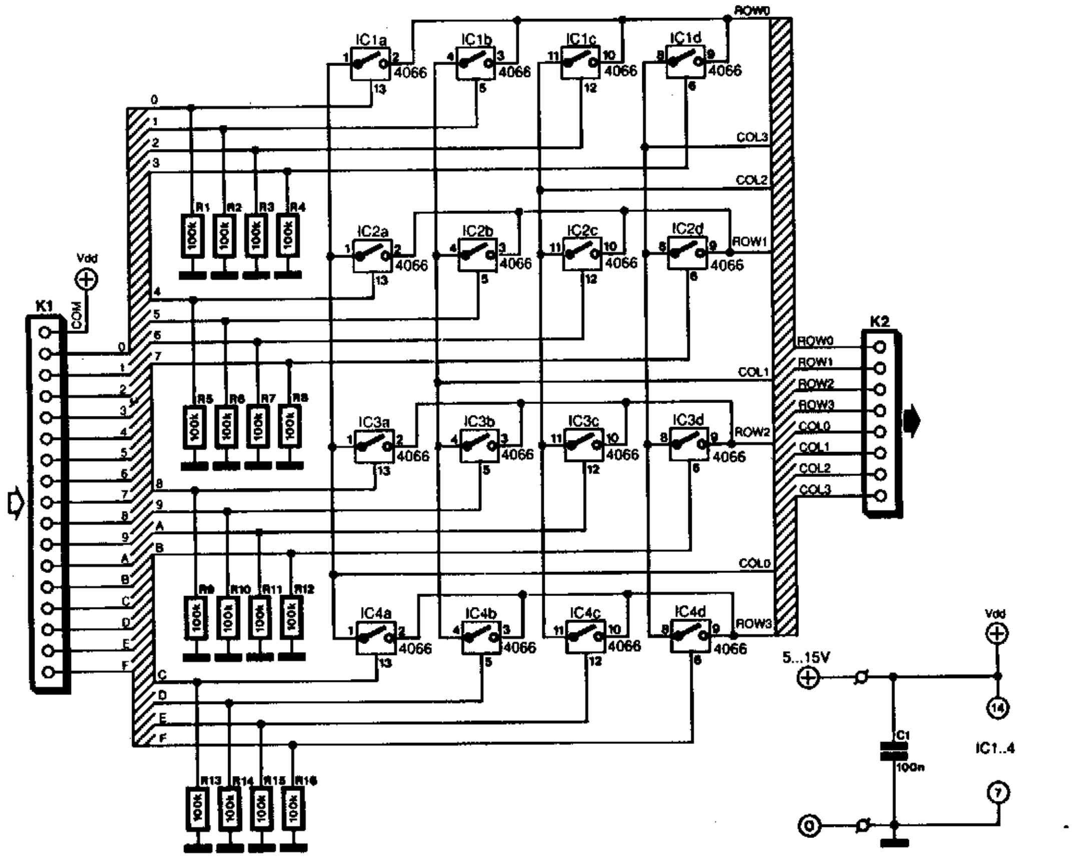
ELECTRONIC DESIGN

Fig. 41-1

Process-control applications often require a monostable multivibrator (one-shot) with a pulse width that can be selected on-the-fly.

This circuit uses two CD4051B analog multiplexers to select the required timing components for the multivibrator, and hence, the pulse width. The multiplexers' address input comes from an 8-bit latch. Bit D6 tells the multivibrator whether to trigger on the leading or trailing edge of the trigger input. Bit D7 determines whether the multivibrator should be in a retriggerable or nonretriggerable mode.

KEYBOARD MATRIX INTERFACE



ELEKTOR ELECTRONICS

Fig. 41-2

Keyboards can be slotted into two categories, at least as far as the manner that the switches are connected is concerned: those with a common connection and those with the switches arranged in a matrix.

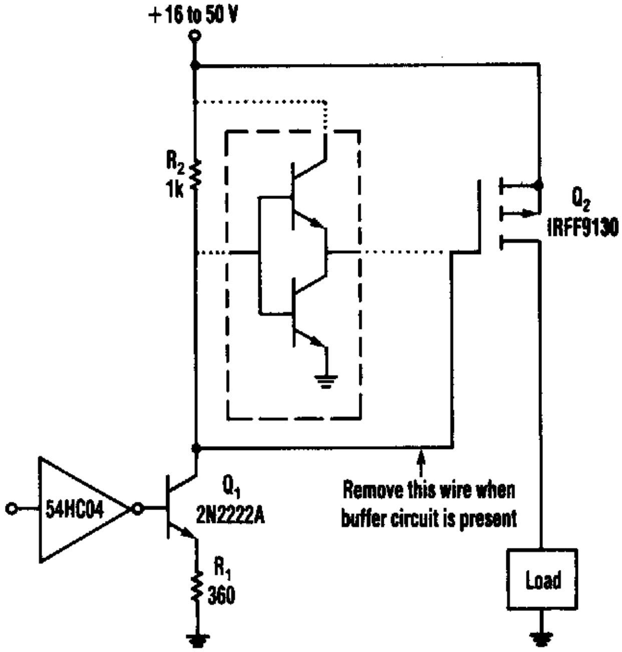
The matrix type has the important advantage that the number of connections is an absolute minimum. Such an arrangement is ideal for ICs; many of these are designed for use with a matrix keyboard.

However, many keyboards are available in job lots, for instance, that apart from a common connection also have a connector for each key. Such keyboards can be connected to ICs that require a matrix type with the aid of a number of electronic switches.

The principle is straightforward: each key of the keyboard controls an electronic switch that is included in a matrix. As an example, the diagram shows a hexadecimal keyboard that is arranged in a 4- × -4 matrix. Each of the electronic switches is held in the open position by a pull-down resistor.

The current drawn by the circuit is very small and is determined mainly by the value of the pull-down resistors and the number of keys being pressed. The CMOS switches draw virtually no current.

LOW-LEVEL POWER FET DRIVER METHOD



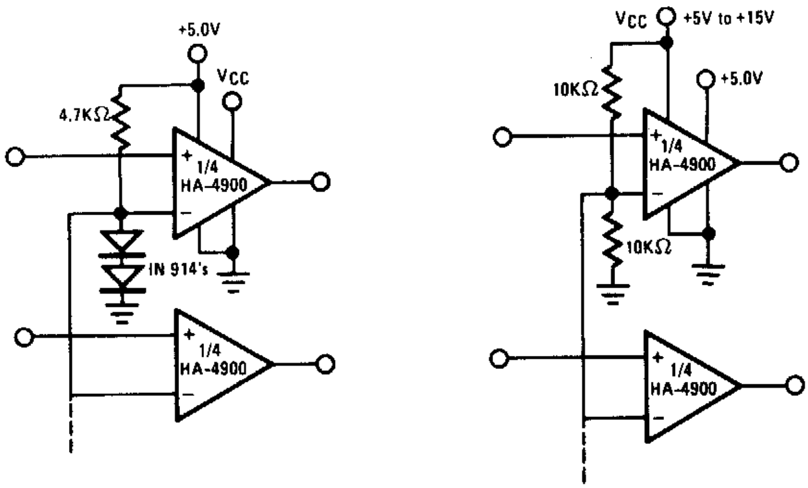
ELECTRONIC DESIGN

Fig. 41-3

This circuit operates from a 16- to 50-V supply. Adding the buffer circuit (within the dashed lines) offers 100-ns switching times. Otherwise, the circuit switches in 1 μs.

Q1 and R1 form a switched current source of about 12 mA. The current flows through R2, which supplies 12 V to the FET. The circuit works well over a wide range of supply voltages. Furthermore, it switches smoothly in the presence of large ripple and noise on the supply. The switching time (about 1 μs) can be reduced considerably by lowering the values of R_1 and R_2 at the expense of higher power dissipations in the resistors and Q1. Alternatively, a buffer circuit can be added to produce switching times of 100 ns without generating significant power dissipation.

LOGIC-LEVEL TRANSLATORS



TTL TO CMOS　　　　CMOS TO TTL

HARRIS　　　　*Fig. 41-4*

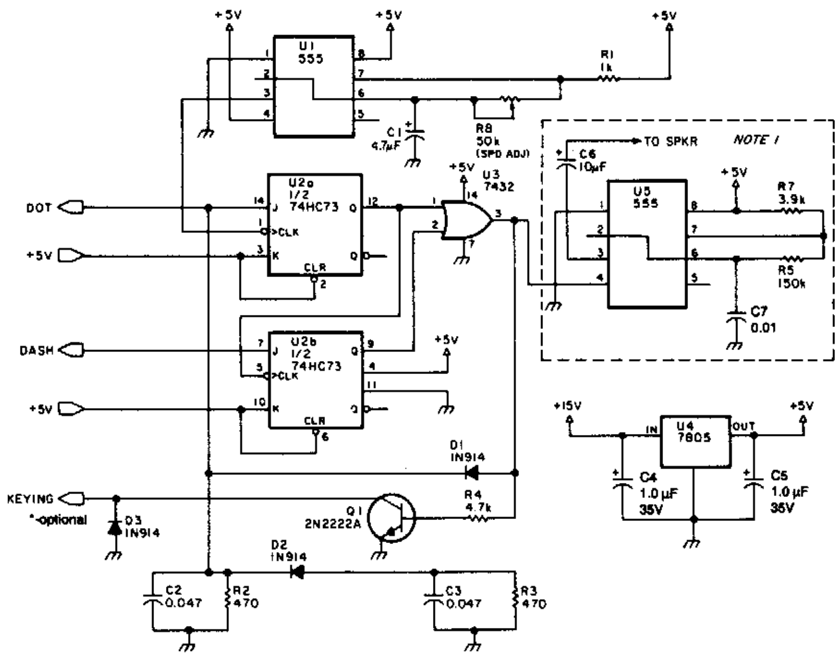
The HA-4900 series comparators can be used as versatile logic interface devices, as shown in these circuits. Negative logic devices can also be interfaced with appropriate supply connections. If separate supplies are used for $V-$ and $V_{LOGIC}-$, these logic-level translators will tolerate several volts of ground-line differential noise.

42

Keying Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 668. The figure number in the box of each circuit correlates to the entry in the Sources section.

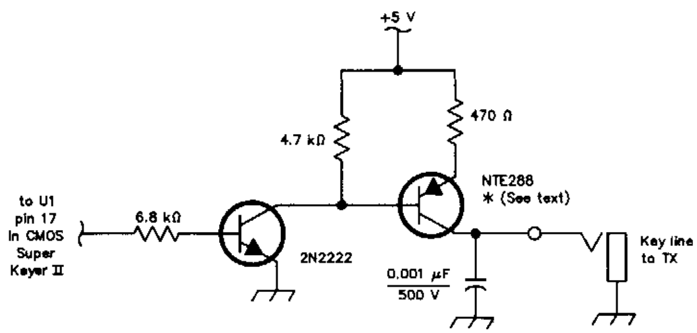
CW KEYER



73 AMATEUR RADIO ***Fig. 42-1***

This electronic keyer uses four ICs (five, if the optional sidetone oscillator is desired) and operates from dc sources of 9 to 15 V. A 2N2222 is used as a keying transistor. If isolators or more power handling ability is desired, a 6-V relay can be keyed with the 2N2222 and the relay in turn can be used to key the transmitter.

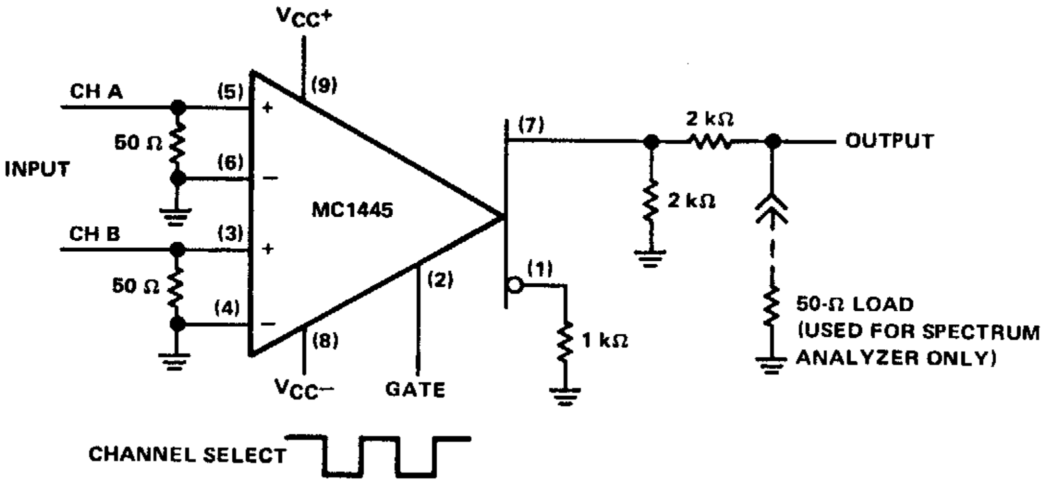
TRANSMITTER NEGATIVE KEY LINE KEYER



QST ***Fig. 42-2***

Using an NTE288 (or ECG288, GE223, or SK3434), this circuit can key a negative line up to −300 V maximum. Do not use this circuit to key a vacuum-tube amplifier that draws grid current because the keying transistor might be damaged under these conditions.

FREQUENCY SHIFT KEYER



TEXAS INSTRUMENTS

Fig. 42-3

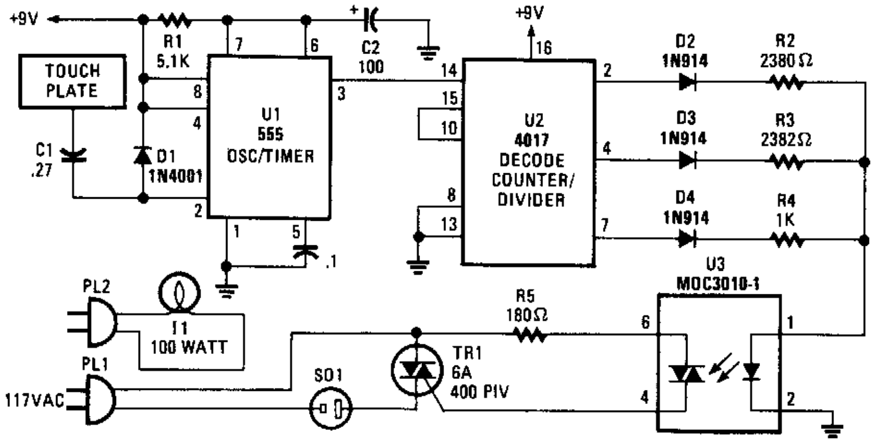
Apply a signal to each differential amplifier input pair. When the gate voltage is changed from one extreme to the other, the output can be switched alternately between the two input signals. When the gate level is high (1.5 V), a signal applied between pins 5 and 6 (channel A) will be passed and a signal applied between pins 3 and 4 (channel B) will be suppressed. In this manner, a binary-to-frequency conversion is obtained that is directly related to the binary sequence, which is driving the gate input (pin 2).

43

Light-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 668. The figure number in the box of each circuit correlates to the entry in the Sources section.

THREE-WAY TOUCH LAMP



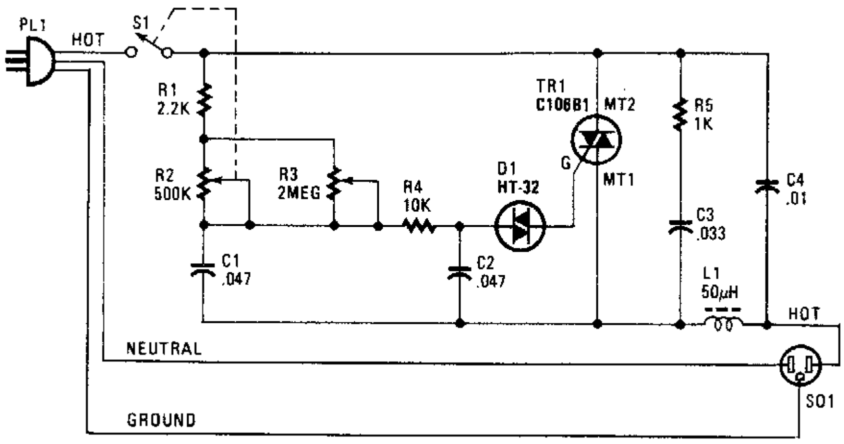
POPULAR ELECTRONICS

Fig. 43-1

A three-way switch to control a lamp (off-dim-bright, etc.) uses an NE555 timer to generate a one-second pulse, triggered by ambient ac fields that are picked up by the human body. C1 and D1 form an input network. U2 is a decode counter/divider and drives one of 10 outputs (three are used). The logic outputs drive various resistors in series with the LED in the optocoupler. The optocoupler controls a triac that is in series with a load (lamp, etc.).

By reconfiguring the outputs of U2, more than three brightness levels can be obtained, up to 10. An IN914 and resistor will be required for each output.

LIGHT DIMMER/SPEED CONTROL

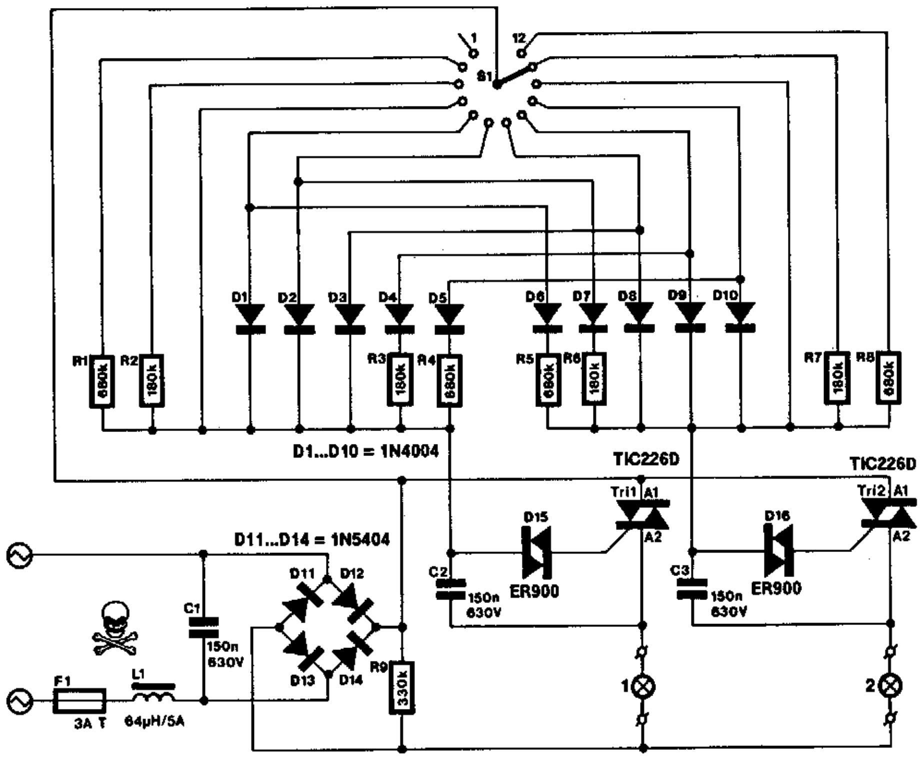


POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

Fig. 43-2

A phase-controlled triac (HT-32) circuit provides control of effective voltage at load. Do not omit L1 and C4 because they are for RFI suppression. The maximum load is about 500 W. WARNING: 120 Vac is present on this circuit—provide adequate insulation and construction techniques.

FOUR-QUADRANT DIMMER



Switch position	Brightness	
	Group A	Group B
1	0	0
2	1/3	0
3	2/3	0
4	1	0
5	1	1/3
6	1	2/3
7	1	1
8	2/3	1
9	1/3	1
10	0	1
11	0	2/3
12	0	1/3

ELEKTOR ELECTRONICS

Fig. 34-3

This very special mains-operated dimmer for domestic or industrial lights is not available in proprietary form: it enables brightness control of two groups of lights in one operation. The possible combinations of brightness are shown in the table. It will be clear that it is not possible to obtain continuous brightness control in the two groups. Instead, the circuit affords the setting of four states of brightness in either group: full on, fully dimmed, 1/3 on, and 2/3 on.

Both sections of the circuit operate on the well-known principle of the triac being switched from the blocking state to the conducting state with the aid of an RC network and a diode. The RC network provides the necessary phase shift and determines when the triac is switched. The rotary switch selects the resistor in a given network, and thus the brightness of the relevant group of lights. No resistor means that the group is off; a short-circuit gives maximum brightness, and resistors of 10 kΩ and 18 kΩ produce intermediate brightness. The diodes prevent the groups from affecting one another.

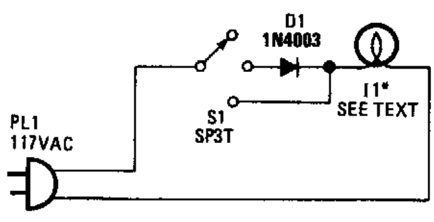
The 64-μH choke (L1) and the 150 nF capacitor across the bridge rectifier prevent the dimmer causing interference in other equipment connected to the mains.

If the triacs are fitted on a heatsink that is rated at 12° K/W, up to 500 W per group can be controlled. It is, of course, essential that the enclosure in which the dimmer is fitted provides ample cooling. A fair number of slots or holes in it are, therefore, essential; these should not permit the circuit elements to be touched.

The switch should have a nonmetallic spindle: this is not only safer than a metallic one, but it also enables the easy removal of the end-notch so that the switch can be rotated continuously, instead of having to be returned to the first stop every time it is operated.

The mains on/off switch S2 should be fitted with a built-in ON indicator bulb, which shows at a glance whether the circuit is on, even though S1 might be in the OFF position. Finally, remember that this circuit carries mains voltage in many places: good workmanship and insulation are, therefore, of the utmost importance.

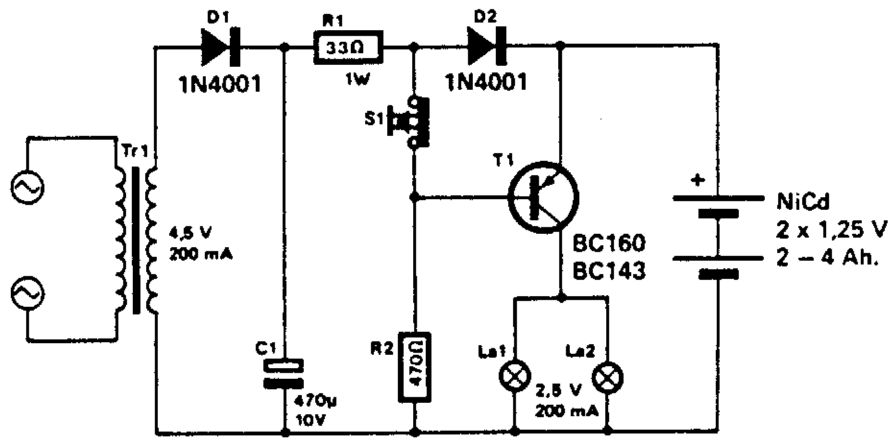
LIGHT DIMMER



POPULAR ELECTRONICS *Fig. 43-4*

Lamp I1 is a household lamp. When the switch is in the center position, the lamp is operated on half-wave rectified ac; the effective voltage the lamp sees is less, which dims it. I1 can be a lamp up to 200 W or 50 rated at 120 or 240 V, and D1 should be a 200-V PIV or better diode (400 PIV for 240-V operation).

AUTOMATIC EMERGENCY LIGHTING UNIT

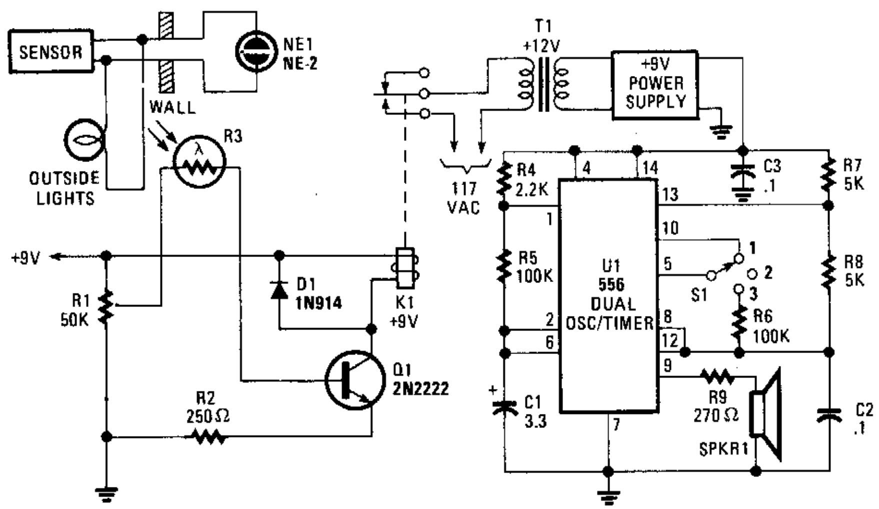


ELEKTOR ELECTRONICS

Fig. 43-5

This unit uses a Nicad battery to provide power to an emergency lighting setup. When power fails, T1 becomes forward biased, which lights L1 and L2. The batteries are normally kept charged. When power is on, T1 is cut off and it keeps the lamps extinguished.

LIGHTS-ON SENSOR

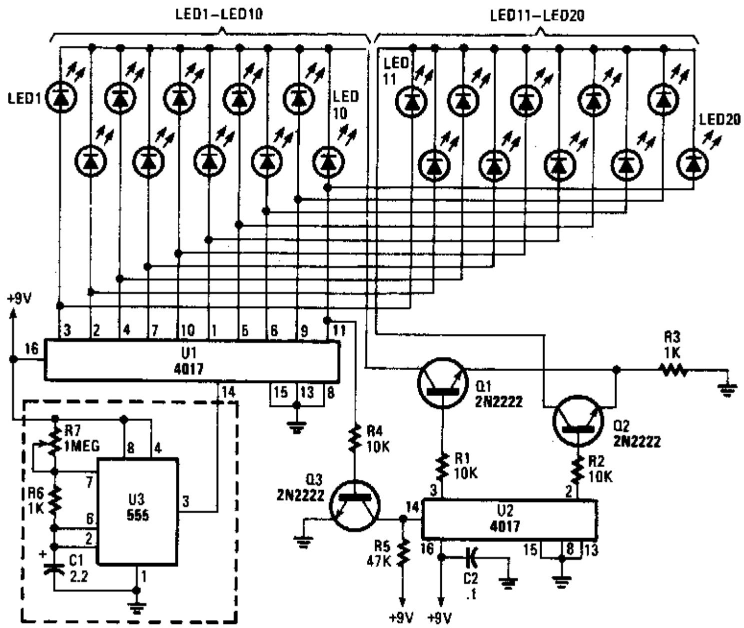


POPULAR ELECTRONICS

Fig. 43-6

Remote monitoring of a light source is possible with this circuit. Photocell R3 activates Q1 and relay K1. U1 is a tone generator that drives a small speaker.

LIGHT CHASER I

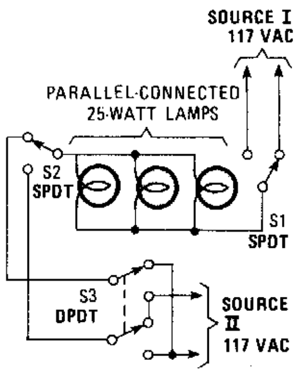


POPULAR ELECTRONICS

Fig. 43-7

Up to 100 lights, LEDs, or optocoupler triac circuits can be sequentially activated by this circuit. One (U1) 4017 decode counter sequences 10 LEDs whose common anode is returned through a second (U2) CD4017, which counts at one-tenth of the rate. The flash rate is controlled by U3, a clock circuit, with a 555 timer.

3-WAY LIGHT CONTROL

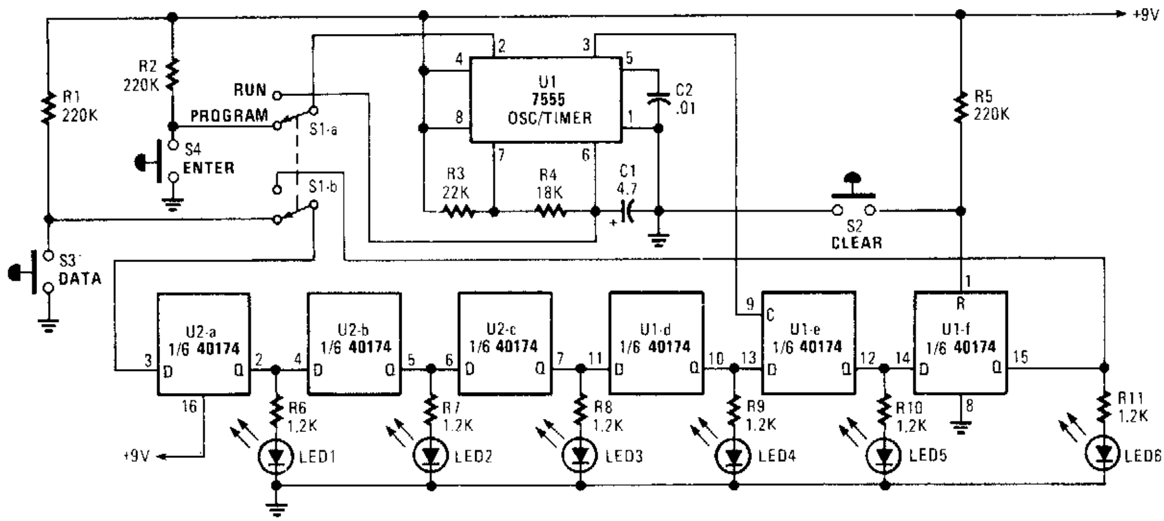


POPULAR ELECTRONICS

Fig. 43-8

This hookup is useful in some house wiring situations, where only two wires are available between switches, rather than the usual 3-way setup where 3 wires are required. S1 and S2 are ordinary three-way switches and S3, a DPDT switch, is commonly available as a four-way switch at hardware stores.

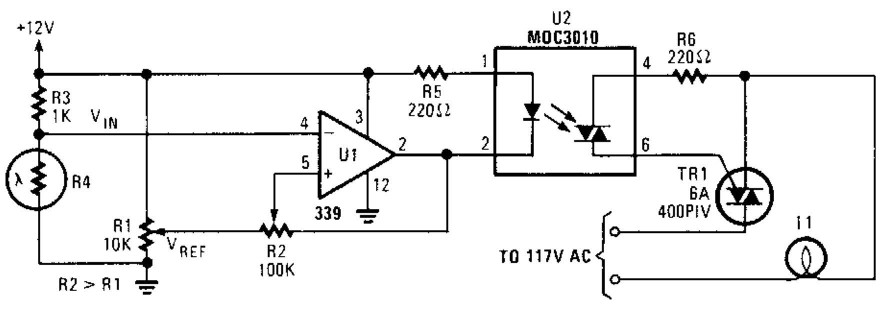
LIGHT CHASER II



POPULAR ELECTRONICS ***Fig. 43-9***

Up to six lights can be sequentially flashed using this circuit. LED1 through LED6 can be replaced by optocouplers (MOC3010, etc.) to control 120-Vac loads via triacs. U1 generates pulses that clock the shift register mode up of the six D flip-flops in the CD 40174. By S1A – B, the register can be programmed either ON or OFF (low or high) and then switched to run in the programmed sequence. S2 clears the program.

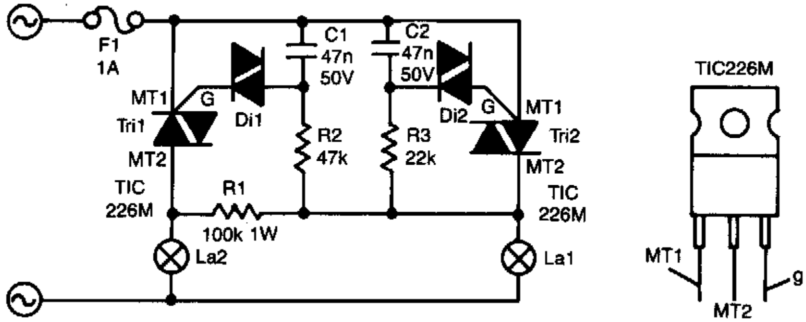
LIGHT CONTROLLER



POPULAR ELECTRONICS ***Fig. 43-10***

A photocell drives U1, a comparator, which controls optocoupler U2. A 6A Triac is used to switch an ac load, such as a lamp, etc.

"AUTOMATIC" LIGHT BULB CHANGER



ELEKTOR ELECTRONICS *Fig. 43-11*

The circuit presented here guarantees that if bulb La1 "gives up the ghost," bulb La2 will take over its task. In series with La1 is triac Tri2. Resistor R3 and C2 form a delay network. As soon as the voltage across C2 rises above about 30 V, diac (gateless triac) D2 is switched on, which causes Tri2 to conduct so that La1 lights.

The control circuit of La2 is parallel to that of La1, but because R2/C1 has twice the delay of R3/C2, Tri1 will not be triggered when Tri2 conducts; C1 discharges so that Tri1 cannot be triggered.

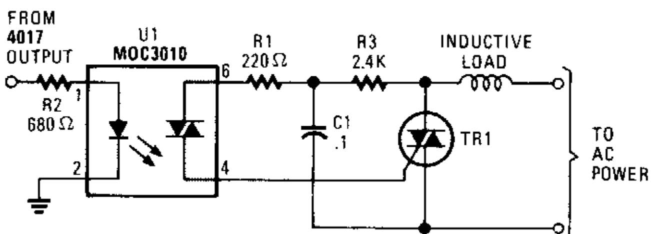
When, however, La1 is open-circuited, a voltage is across both RC networks via La2 and R1. Again, Tri2 will be triggered first, but because the current is smaller than its holding current, it will cease to conduct almost immediately. Capacitor C1 will then continue to charge and after a little while Tri1 is switched on.

Because the time constant for La2 is somewhat longer than that for La1, La2 will always be slightly less bright than La1. It is, of course, possible to give La2 a slightly higher wattage than La1 to ensure equal brightness.

Without heatsinks, the triacs can handle up to 100 W each; with heatsinks, powers of up to 1 000 W can be accommodated. It is not recommended to use bulbs with a wattage below 25 W, because these can flicker.

The triacs can be any type that can handle at least 400 V at no less than 5 A. The M types used in the prototype can handle 600 V at 5 A.

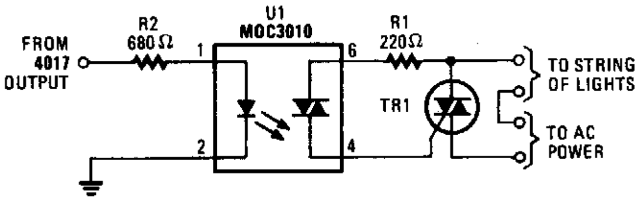
INDUCTIVE LOAD TRIAC SWITCH



POPULAR ELECTRONICS *Fig. 43-12*

An additional resistor and capacitor enable control of an inductive load, such as a small blower motor, fluorescent lamp, etc.

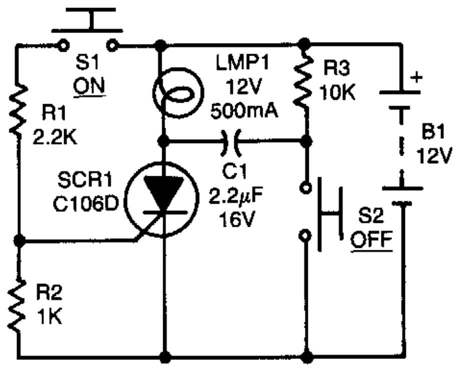
CHRISTMAS LIGHT DRIVER



POPULAR ELECTRONICS *Fig. 43-13*

This circuit will enable a CMOS logic chip, such as a 4017 decode driver, to control a string of Christmas lights or other lighting. The triac should be rated at 200 V and 3 A or higher. The 4017 should be powered from at least 10 V to ensure adequate drive to the optoisolator.

SCR CAPACITOR TURN-OFF CIRCUIT



RADIO-ELECTRONICS *Fig. 43-14*

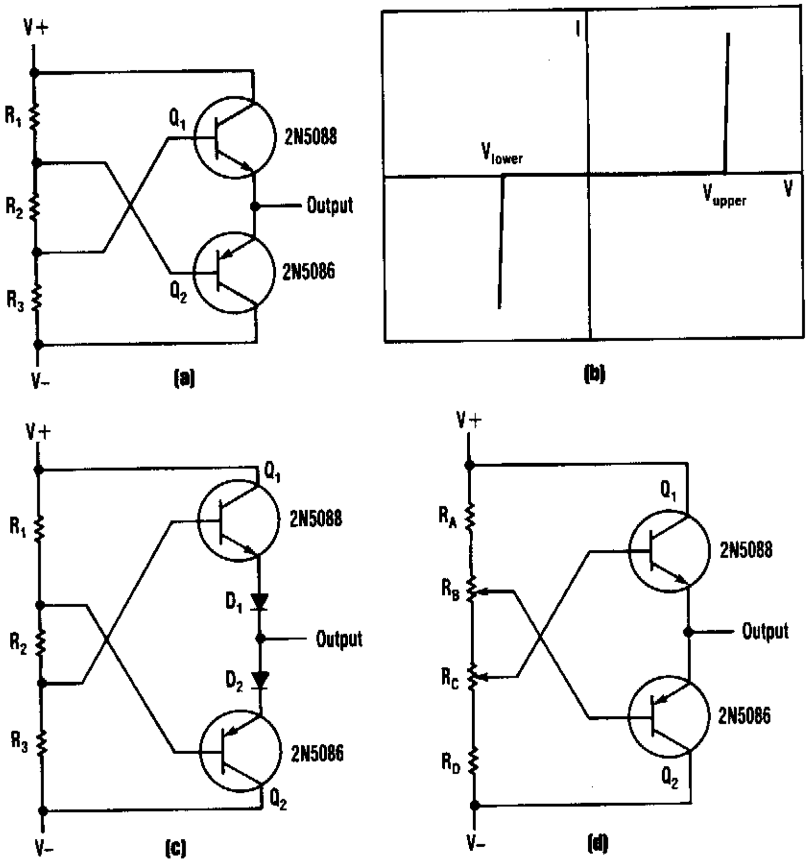
After the SCR turns on, C1 charges up to almost the full supply voltage via R3 and the anode of the SCR. When S2 is subsequently closed, it clamps the positive end of C1 to ground, and the charge on C1 forces the anode of the SCR to swing negative momentarily, thereby reverse-biasing the SCR and causing it to turn off. The capacitor's charge bleeds away rapidly, but it has to hold the SCR's anode negative for only a few μs to ensure turn-off. C1 must be a nonpolarized type.

44

Limiter

The sources of the following circuits are contained in the Sources section, which begins on page 668. The figure number in the box of each circuit correlates to the entry in the Sources section.

ADJUSTABLE VOLTAGE LIMITER



ELECTRONIC DESIGN

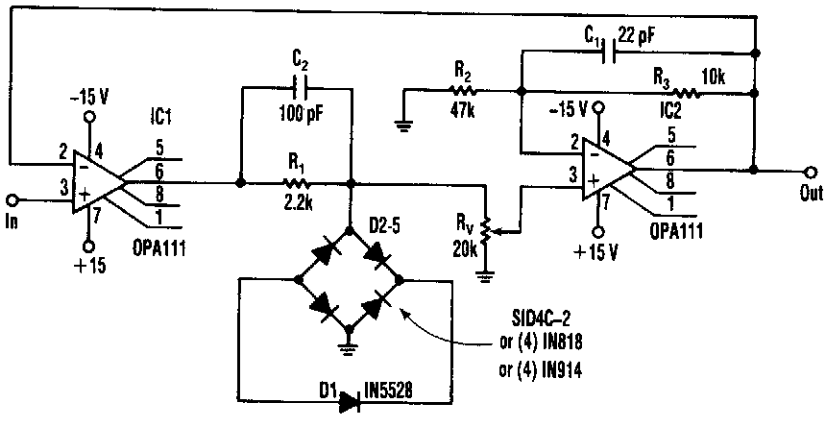
Fig. 44-1

This bipolar voltage clipper can be built with two transistors, two resistors, and two potentiometers. Notice that the maximum p-p range must be ≤ BV_{ebo} of the transistors used. The design equations are:

$$V_{\text{upper}} = V_- + \frac{(V_+ - V_-)\,(R_2 + R_3)}{(R_1 + R_2 + R_3\ \ )} + V_{be}(Q_2)$$

$$V_{\text{lower}} = V_- + \frac{(V_+ - V_-)\,(R_3)}{(R_1 + R_2 + R_3)} - V_{be}(Q_1)$$

ONE-ZENER PRECISE LIMITER



ELECTRONIC DESIGN

Fig. 44-2

A limiter circuit that requires matched zener diodes can instead use one zener with a full-wave diode bridge. The circuit's two limits are nearly equal when determined by the same zener—only two pairs of forward diodes need to be matched. For best results, an integrated quad of diodes can be used. But, after testing the circuit, four single controlled-drop diodes and four ordinary diodes gave about the same accuracy (better than 0.5%).

Because the limiting level can be adjusted, zener tolerance can be adjusted out. Gain stability can be optimized by connecting the inverting input to the first op amp to the output of the second to make the circuit inherently unity-gain.

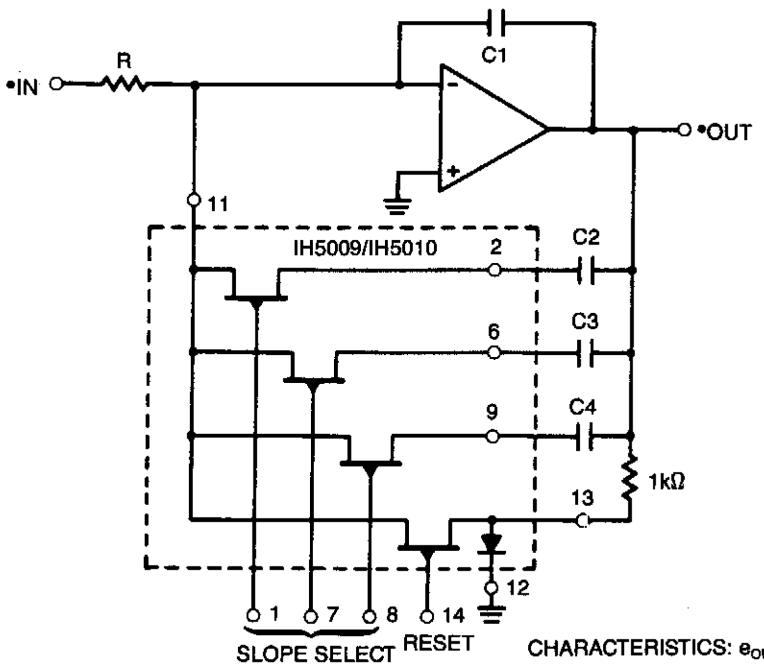
The zener voltage must be increased to 8.2 V to compensate for the two diode drops. Placing small capacitors across the resistors in the loop stabilized the circuit adequately and response is orders of magnitude faster than conventional circuits. Moreover, it's limited primarily by the op amp's slew rate.

45

Mathematical Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 668. The figure number in the box of each circuit correlates to the entry in the Sources section.

PROGRAMMABLE SLOPE INTEGRATOR



CHARACTERISTICS: $e_{OUT} = -\frac{1}{RC}\int e_{IN}\,dt$

INTERSIL

Fig. 45-1

By using analog switch IH5009/IH5010 to select various capacitors, a variable slope integrator can be had. If $C_3 = 2(C_2)$ and $C_4 = 4(C_2)$, seven different slopes can be obtained if binary information is fed to pins 1, 7, and 8 of the analog switch.

FAST BINARY ADDING CIRCUITS

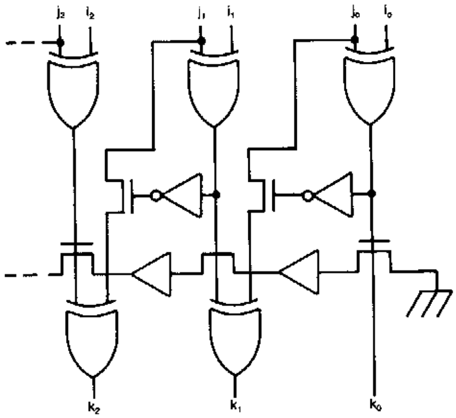


Fig. 45-2(a)

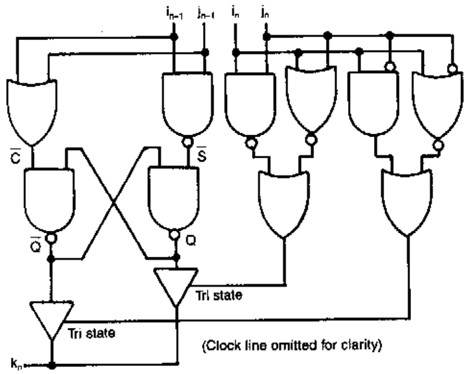


Fig. 45-2(b)

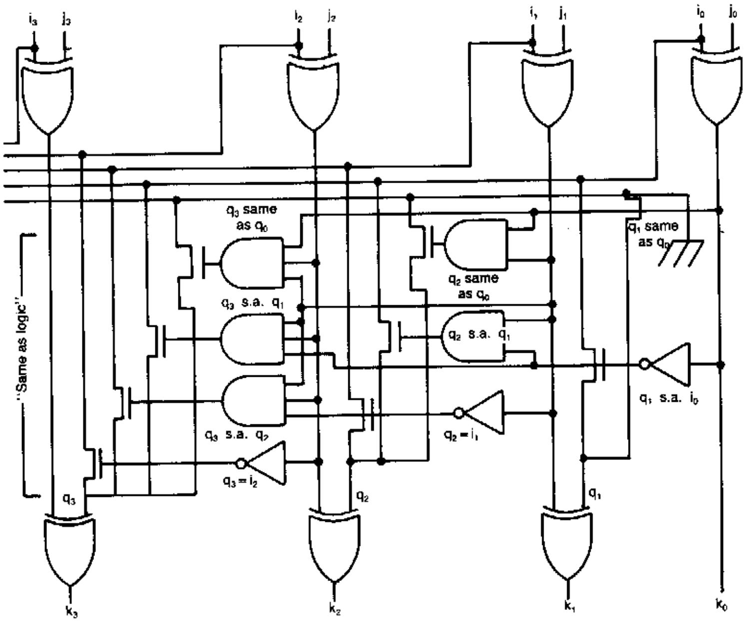


Fig. 45-2(c)

Some circuits that add binary numbers have problems with time delay caused by carry propagation. This has been partially solved by the carry look-ahead adder. However, because of the complexity of this scheme, the carry look-ahead logic usually covers no more than 4 bits, and a ripple carry is implemented between the carry look-ahead blocks.

The Daniels Adder avoids these problems by presenting a scheme where carry bits are not used at all in the process of binary addition. It is based on recognition patterns, which exist with the binary addition truth table.

The addition is described by the following two sets of equations:

$$\text{if } i_{n-1}=j_{n-1} \qquad q_n=i_{n-1}$$

$$\text{if } i_{n-1}=j_{n-1} \qquad q_n=q_{n-1}$$

$$\text{if } i_n=j_n \qquad k_n=q_n$$

$$\text{if } i_n=j_n \qquad k_n=q_n$$

with the boundary condition that $q_{-1}=0$, where i_n, j_n, and k_n are the bit of binary weight 2^n (n^{th} bit) of the addend, summand, and sum respectively, q_n is an intermediate variable and q_n is the inverse of q_n.

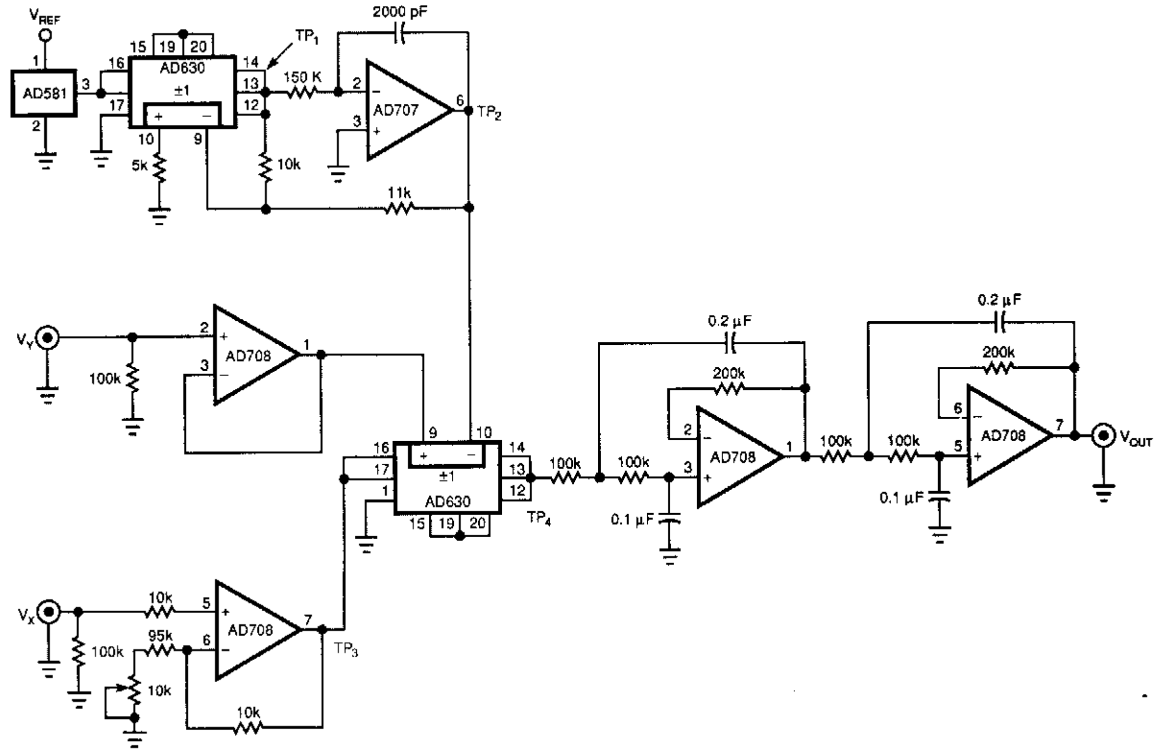
The value of the sum is (depending upon i_n and j_n) either the same as or the inverse of (depending upon i_{n-1} and j_{n-1} a0, a1, or the inverse of the $(n-1)^{\text{th}}$ bit of the sum. Figure 45-1(a) shows the logic diagram of the ripple through implementation of the adder.

Because each stage calculates whether its value of the intermediate variable q_n is the same as the previous stage's value (q_{n-1}) in parallel, it is possible to devise simple "same as" logic that does not have the complexity drawback of carry look-ahead logic and can be carried over any number of bits (Fig. 45-1(b)). A 32-bit adder built in this way will result in 11-gate delays (no gate having more than 4 inputs).

Especially compact and efficient is the pipelined implementation (Fig. 45-1(c)), which can produce the sum at a rate of 3-gate delays/bit.

The high-speed adder circuits can be used on gate arrays or full-custom ICs to implement fast calculation of addresses or data values. Because of their compact nature, they also use less space on the silicon than conventional adders do.

MULTIPLYING PRECISE COMMUTATING AMP



EDN

Fig. 45-3

By using a pulse-width-height modulation technique, this circuit implements a 0.015%-accurate multiplier. The circuit's output equals $V_X V_Y/10$. An AD581 voltage reference, an AD630 commutating amplifier, and an integrator comprising an AD707 op amp, 2000-pF capacitor, and 150-kΩ resistor first generate a precision triangle wave. For a given state of the AD630's output ($+V_{REF}$ at TP_1, for example) the integrator ramps until its output reaches −11 V. Then, TP_1 changes state and the integrator begins ramping toward +11 V. The triangle wave's period is $4.4RC$ or 1.32 ms, where R and C are the values of the integrator components.

The circuit uses a second AD630, driven by the variable V_X to compare the triangle waveform at TP_2 to the signal at V_Y. The duty cycle, T_1+T_2, at the output of this second commutating amplifier is:

$$T_1=\frac{2RC(11-V_Y)}{10}$$

and

$$T_2=\frac{2RC(11+V_Y)}{10}$$

During T_1, the voltage at TP_4 equals $-1.1V_X$. During the remaining period, T_2, the pulse height will equal $+1.1V_X$. V_{OUT} is the average, obtained by low-pass filtering, of this T_1 and T_2 combined waveform and equals:

$$V_o=\frac{-1.1V_XT_1+1.1V_XT_2}{T_1+T_2}=\frac{V_XV_Y}{10}$$

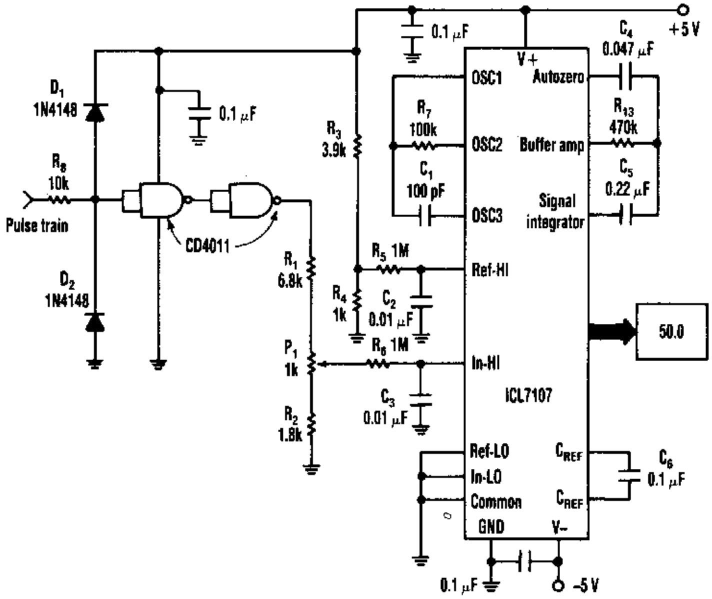
You can use a higher bandwidth filter and a higher carrier frequency to build a faster multiplier.

46

Measuring and Test Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 669. The figure number in the box of each circuit correlates to the entry in the Sources section.

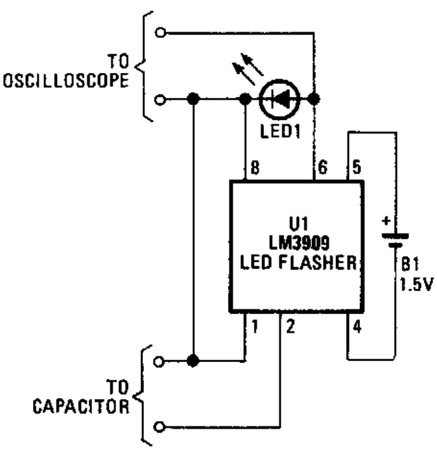
DUTY-CYCLE MEASURER



ELECTRONIC DESIGN **Fig. 46-1**

An Intersil 7101 3½-digit A/D converter is used to display the duty cycle of a pulse train as a percentage. The frequency range of this circuit is 100 Hz to 250 kHz. The CMOS gates convert the pulse train to constant amplitude. This amplitude is then compared to a reference of 1 V, derived from R3 and R4. P1 is for calibration.

SIMPLE CAPACITOR TESTER



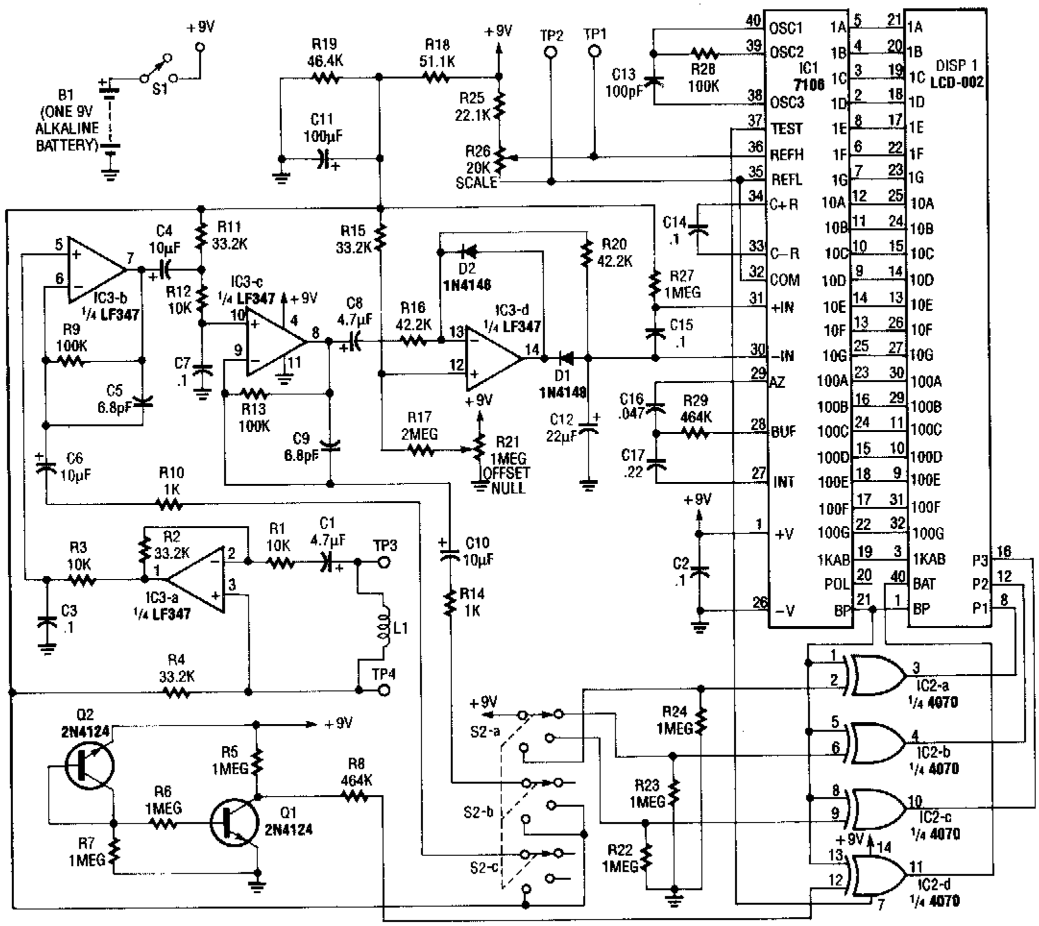
POPULAR ELECTRONICS **Fig. 46-2**

An LM3909 LED flasher is used as an oscillator, and the capacitor connected to the terminals determines frequency.

C	f approx
4 700 μF	0.04 Hz
470 μF	0.4 Hz
4.7 μF	4 Hz
0.047 μF	4 kHz
0.004 7 μF	25 kHz
47 μF	660 kHz

The LED can be used to count frequency visually using a stopwatch for large capacitors ($C > 500$ μF).

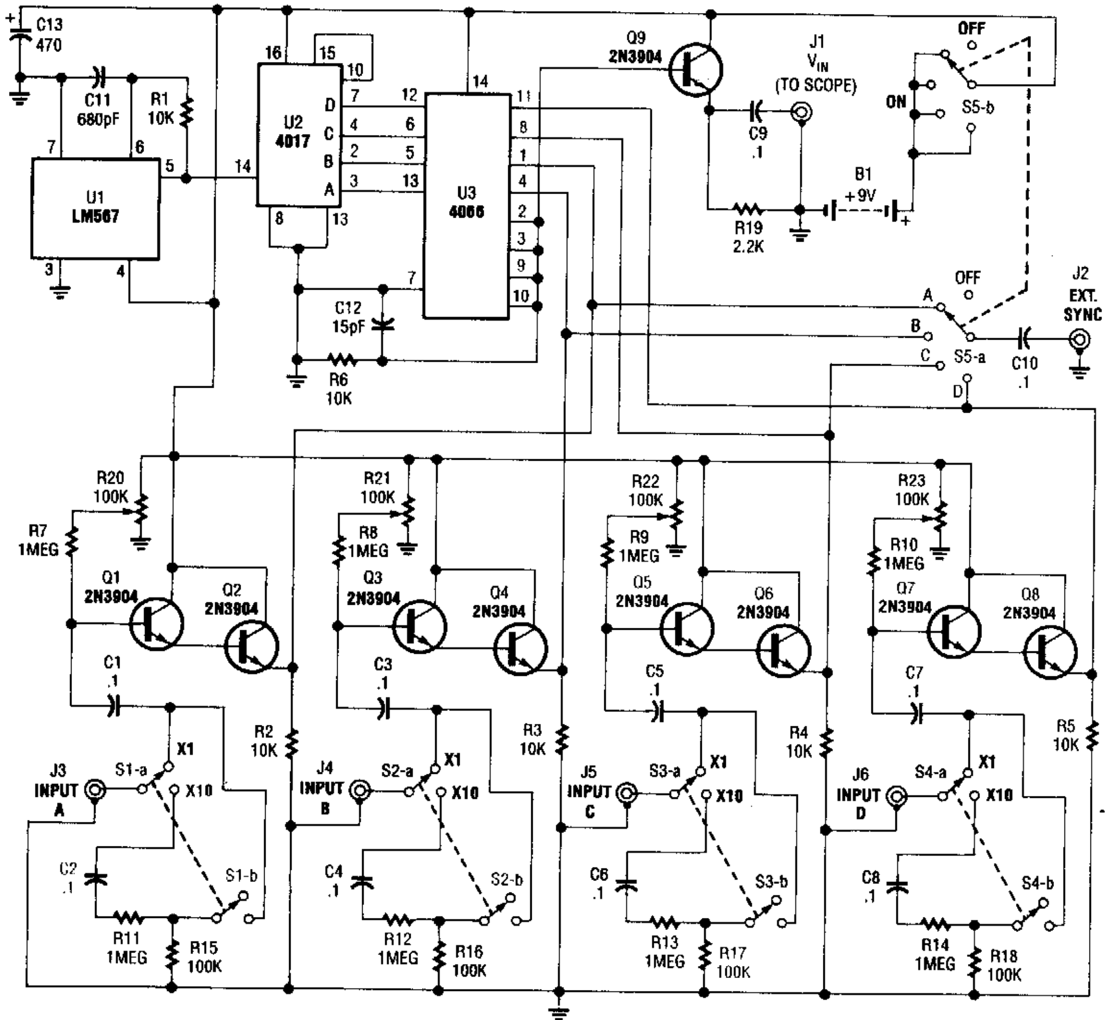
MAGNETIC FIELD METER



RADIO-ELECTRONICS *Fig. 46-3*

Using a pickup coil to drive an amplifier (IC3A-B-C-D), this meter circuit can be directly calibrated in field-intensity units. R3/C3 and R12/C7 establish a frequency roll off that compensates for the pickup-coil sensitivity, and set a 20-kHz cut-off point. S2 is the range-select switch. L1 is an 18-turn 3″ diameter coil. The frequency range is 50 Hz to 20 kHz and the range of measurement is 0.1 to 20 000 microTesiers (μT).

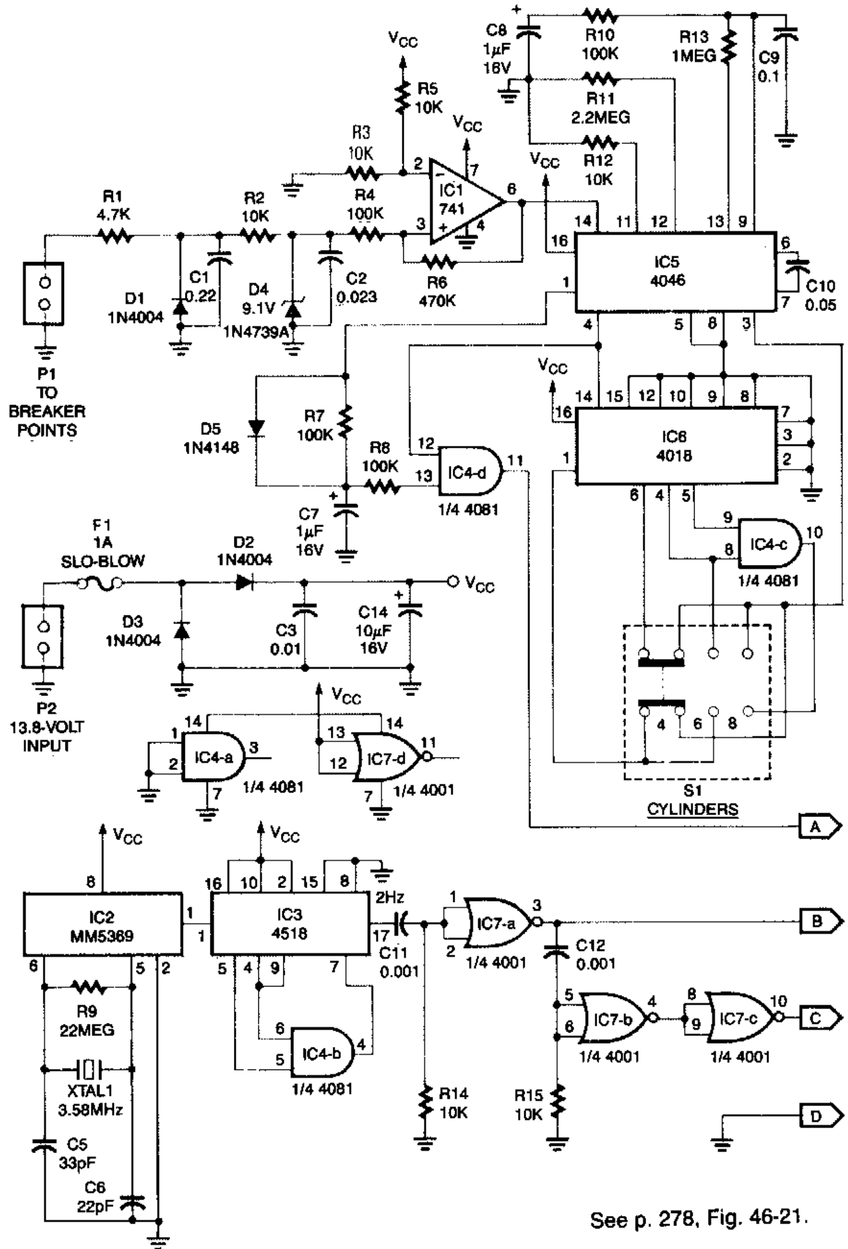
FOUR-TRACE OSCILLOSCOPE ADAPTER



POPULAR ELECTRONICS *Fig. 46-4*

This simple adapter uses an oscillator (567) to drive a counter (U2) and switch (U3) that selects the output of one of four scope preamps (Q1/Q2 through Q7/Q8) and feeds it to buffer Q9 and output jack J1. J2 provides synch to the scope. R20 through R23 are posting controls for channels A through D (J3 through J6). S1A-B through S4A-B are switched attenuators, one for each channel. Switching rate is about 125 kHz. This circuit is useful for adding four-trace operation to inexpensive oscilloscopes. Signal levels of 0 to 20 V can be handled.

DIGITAL TACHOMETER CIRCUITRY



See p. 278, Fig. 46-21.

RADIO-ELECTRONICS

Fig. 46-5

DIGITAL TACHOMETER CIRCUITRY *(Cont.)*

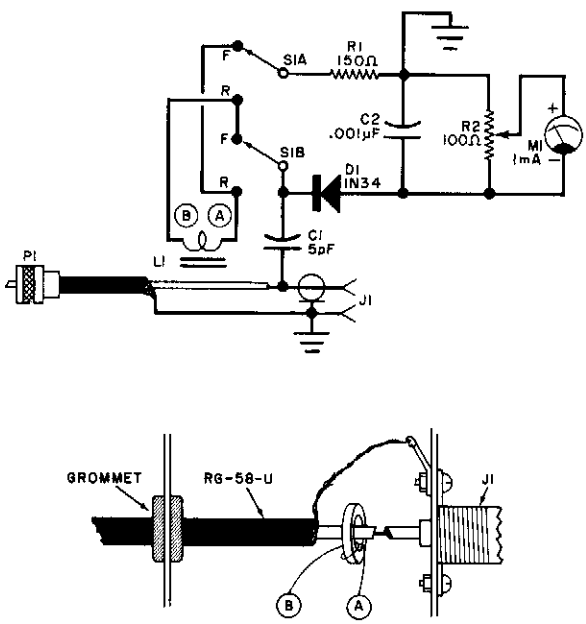
This system can be used with 4-, 6-, or 8-cylinder automobiles. The timebase formed by IC5 is an oscillator that drives counter IC6, which divides by 6, 4, or 3 for 4-, 6-, or 8-cylinder engines, respectively. S1 selects this number. IC5 produces a signal that is phaselocked to this multiple of the ignition system frequency, which in turn depends on engine speed.

$$\text{freq} = \text{rpm} \times \frac{\#\ \text{cylinders (4, 6, or 8)}}{120}\ \text{Hz}$$

IC1 conditions the ignition input at P1 to feed IC5. The output of IC4D, which is the same frequency as the VCO in IC5, is fed to the frequency display.

IC2 generates a 60-Hz signal using a 3.58-MHz reference. IC3 and IC4B divide this by 30 to produce 2 Hz. IC7B/IC7C and C12/R15 produce a delayed 2-Hz signal. These signals are fed to the counter circuit.

SENSITIVE SWR METER



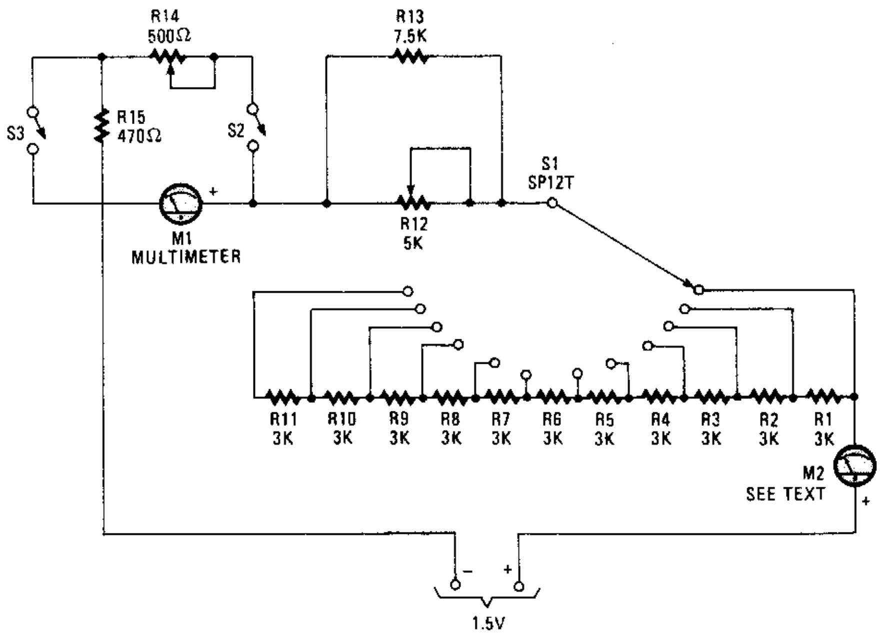
Reflected Meter Reading (% of full scale)	SWR
0	1:1
10	1.22:1
20	1.5:1
30	1.85:1
33.5	2:1
40	2.33:1
50	3:1
60	4:1
66.67	5:1
70	5.66:1
80	9:1
90	19:1
100	∞:1

POPULAR ELECTRONICS

Fig. 46-6

Using a toroidal pickup coil around the center conductor of a coaxial cable, this circuit can be used to measure the SWR of an antenna. L1 is two turns #26 enamelled wire on a Fair-Rite 5963000301 toroidal core.

METER TESTER



POPULAR ELECTRONICS ***Fig. 46-7***

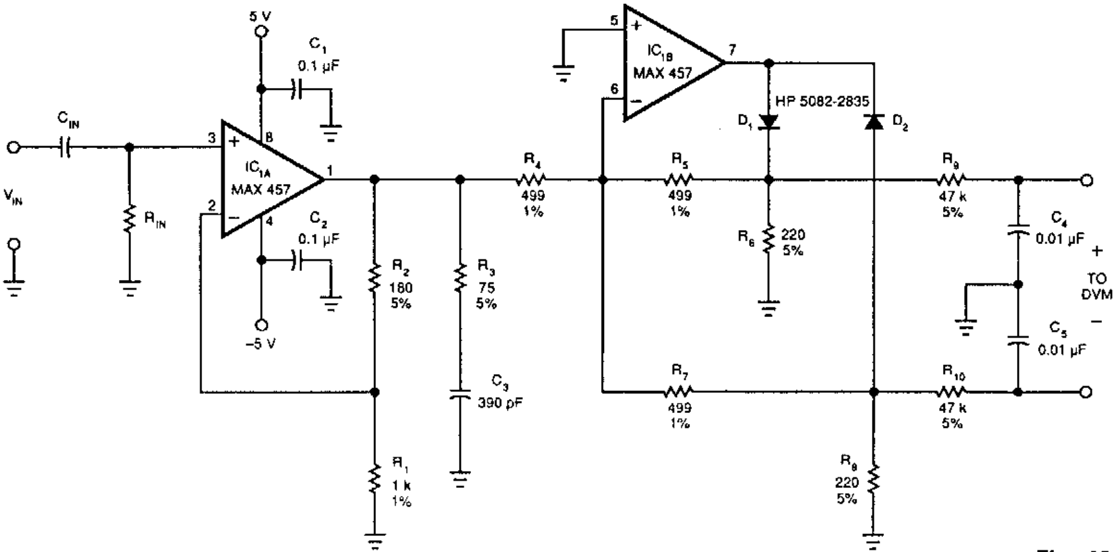
This unit uses switches and resistors to provide a number of current ranges. It allows you to test most of the meters available at surplus outlets, and without damaging the sensitive movements when you have no idea of internal resistance or full-scale current of the unit.

M1 is a multimeter set to measure current, and M2 is the meter-under-test. Starting with S1 set at the maximum resistance and S2 open, decrease the resistance setting of S1, fine tuning with R12, until M2 reads full scale. Then, read M1. It will tell you the full-scale current for the unknown meter. As the meters are connected in series, the same current flows through both.

Now, close S2 and adjust R14 and R15 until M2 reads exactly mid-scale and M1 reads the same current as determined earlier to be the maximum current for M2. Half the current is flowing in M2 and half is going through R14 and R15. The voltage drop is the same across the meter and R14 and R15, because they're in parallel. Thus, the sum of the resistance of R_{14} and R_{15} is the same as the internal resistance of meter M2.

If the internal resistance of M2 is less than 470 Ω, set R14 at maximum resistance and close S3. Readjust R14. Both R14 and R12 should be linear-taper potentiometers.

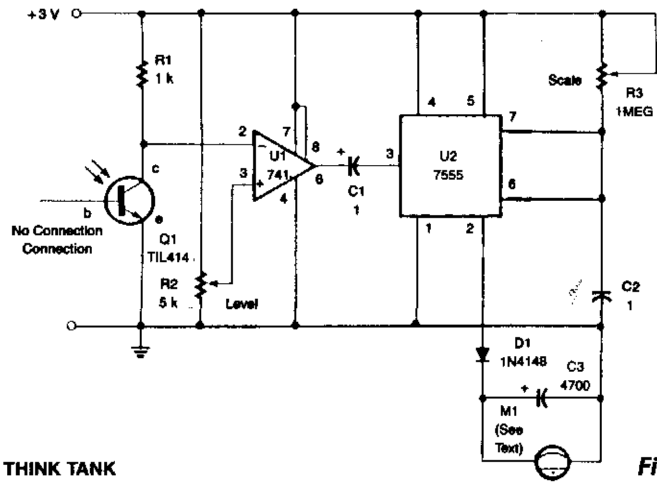
BROADBAND AC ACTIVE RECTIFIER



EDN *Fig. 46-8*

This circuit converts sine waves of up to 1-V rms into an equivalent dc level. It should prove useful as an ac broadband voltmeter. IC1 is an input amplifier that converts level from rms to an equivalent dc level and feeds ICB1. R3 and C3 are stabilizing components. IC2 acts as a full-wave rectifier. R6, R5, and D1 rectify positive levels, R7, R8, and D2 negative-going signals. R9, R10, C4, and C5 are low-pass filters. The output can feed a DVM or another meter.

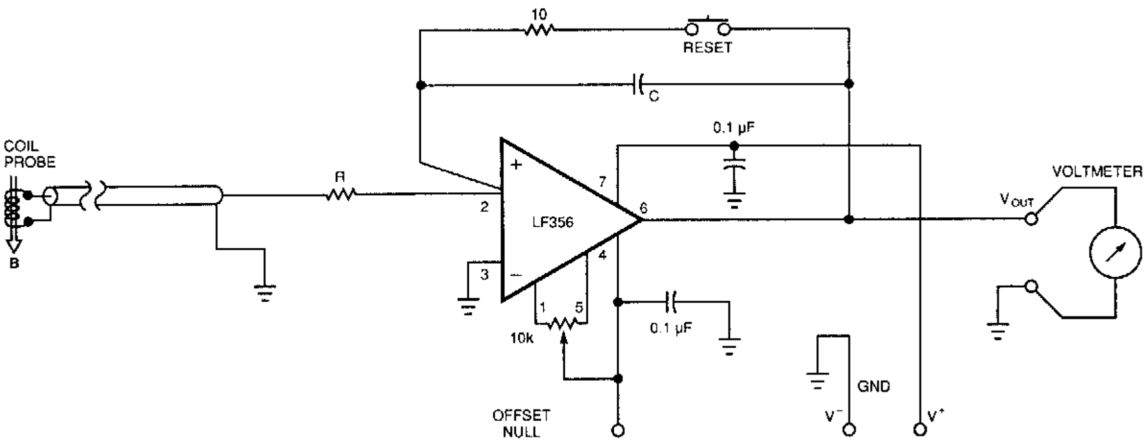
BIKE SPEEDOMETER



WELS' THINK TANK *Fig. 46-9*

A TIL414 photo transistor senses reflection from a spoke-mounted reflector. This generates a pulse and sends it to U1 and U2, a monostable multivibrator, which drives meter M1.

B-FIELD MEASURER



EDN

Fig. 46-10

This circuit develops an output voltage that is proportional to the magnetic induction, B, flowing through its probe's coil. You must size the coil to give a full-scale, 10-V output for your maximum expected magnetic-induction intensity.

For a given value of B (in tesla) and output voltage, V_{OUT}:

$$B = \frac{(R \times C \times V_{OUT})}{A}$$

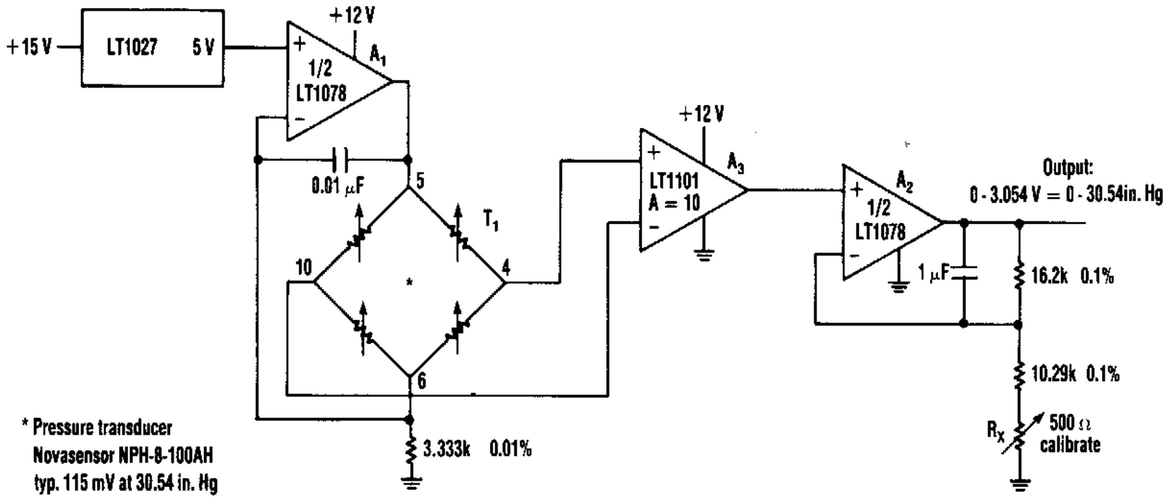
where A is the effective area of your coil in m^2 (A = number of turns × average area of each turn), R is the resistance of the coil and the probe, and C is the value of the capacitor. Notice that C should be a low-leakage polypropylene or Teflon device.

For most practical applications that measure a magnetic field in the air, the coil will be either tiny or very thin. If $R = 1\ k\Omega$, $C = 1\ \mu F$, and the coil is 100 turns with a mean area per turn of 1 cm^2, then the circuit's output will be 1 mV/gauss ($1T = 104G$).

To use the circuit, push the reset button and place the probe in an area that you know is devoid of magnetic fields. Be sure to avoid magnets and iron. Then, put the probe into the field to be measured and read the V_{OUT} with a voltmeter. Finally, calculate the B field's intensity using the equation.

When constructing the instrument, guard the op amp's inputs from undesirable currents at the minus input. For full-scale outputs, use a ±15-V supply for the op amp.

LOW-COST BAROMETER

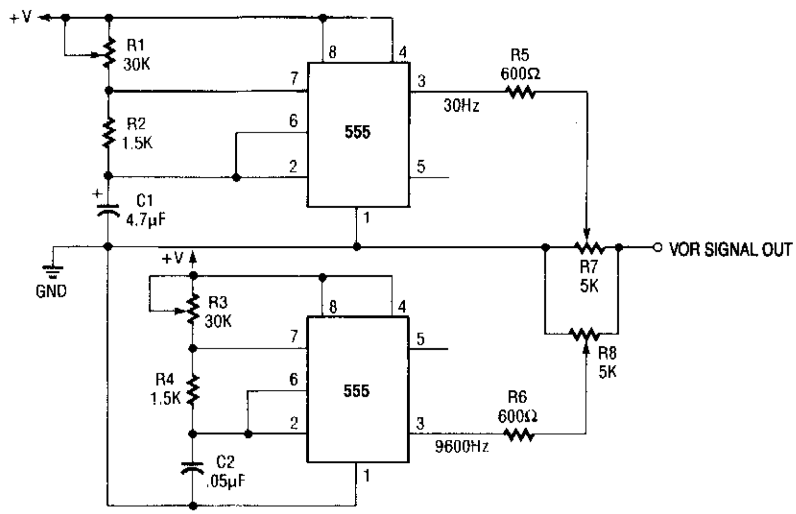


ELECTRONIC DESIGN

Fig. 46-11

Using Linear Technology LT1027 reference and LT1078 op amps, transducer T1 is fed with 1.5 mA. The pressure transducer feeds an amplifier with a gain of 10, then it feeds a voltage follower. Output can either drive an analog meter or a DVM circuit.

VOR SIGNAL SIMULATOR

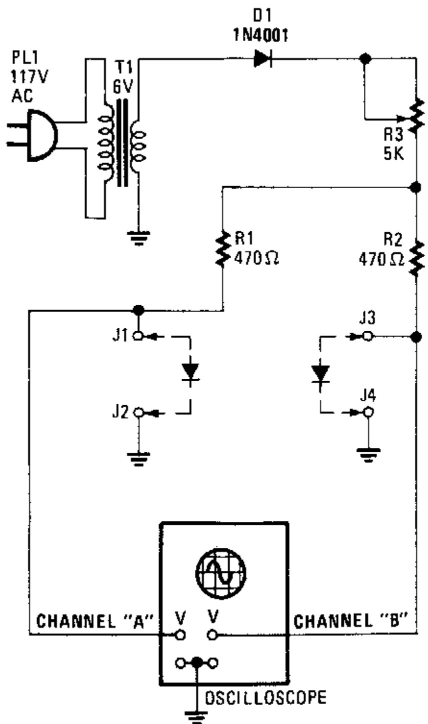


RADIO-ELECTRONICS

Fig. 46-12

An output of 9 960 and 30 Hz at 0.5 Vrms is produced by this VOR (VHF Omni Range) signal simulator, which uses a pair of 555 timers. Alternatively, a 556 (two timers in one package) could be used.

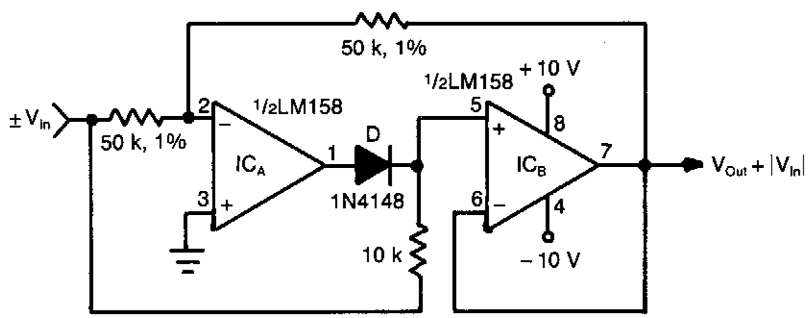
SIMPLE DIODE CURVE TRACER



POPULAR ELECTRONICS *Fig. 46-13*

Suitable for matching diodes or examining *V/I* characteristics of two terminal devices (diodes, etc.), this circuit should be handy for lab use. R_1 and R_2 can be increased in value and a higher voltage transformer can be used for higher voltage test using this principle.

SIMPLE ABSOLUTE VALUE CIRCUIT

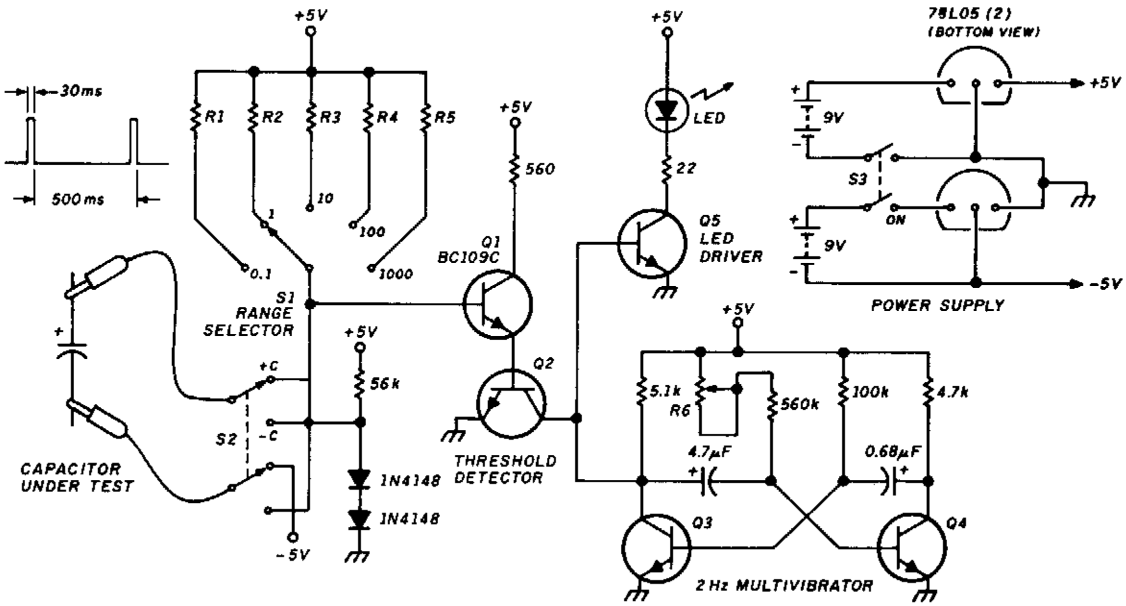


ELECTRONIC DESIGN *Fig. 46-14*

When an input voltage is positive, the output of ICA is negative and diode D does not conduct; hence the output of ICB is positive. On the other hand, when the input is negative, the output of ICA is positive. D will conduct and cause the absolute value, expressed as a positive voltage, to appear on the noninverting input of ICB and on the circuit's output.

The circuit's dynamic range extends from zero to the point at which the op amp saturates. The bandwidth is determined by the characteristics of the diode and the high-frequency performance of the op amp.

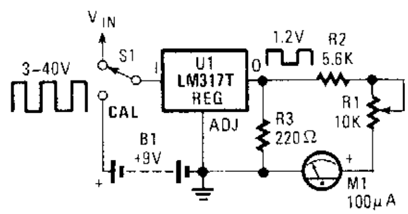
MICROFARAD COUNTER



HAM RADIO ***Fig. 46-15***

This circuit measures capacitance by the time it takes an unknown capacitor to reach 6.32 V (10 V × 63% or 1 *RC* time constant) when charged through resistor R. The LED used as a timebase is pulsed by Q3, Q4, and Q5. By counting seconds (two flashes per second) until threshold detector Q1/Q2 stops the count, you can directly read the number of microfarads. R_1, R_2, R_3, R_4, and R_5 can be any convenient values, such as 1 kΩ, 10 kΩ, 100 kΩ, 1 MΩ, 10 MΩ.

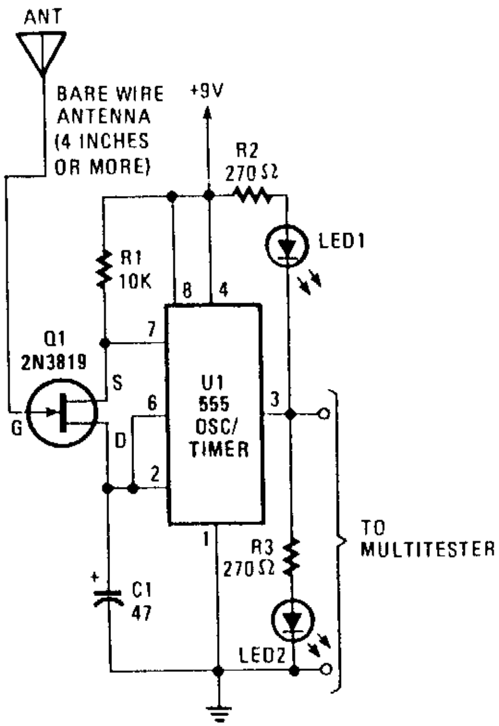
SIMPLE DUTY-CYCLE METER



POPULAR ELECTRONICS ***Fig. 46-16***

Using an LM317T as a precision clipper, this displays the duty cycle of a pulse train from 0 to 100% on a standard 100-mA analog panel meter. This circuit will work up to about 50 kHz.

STATIC DETECTOR

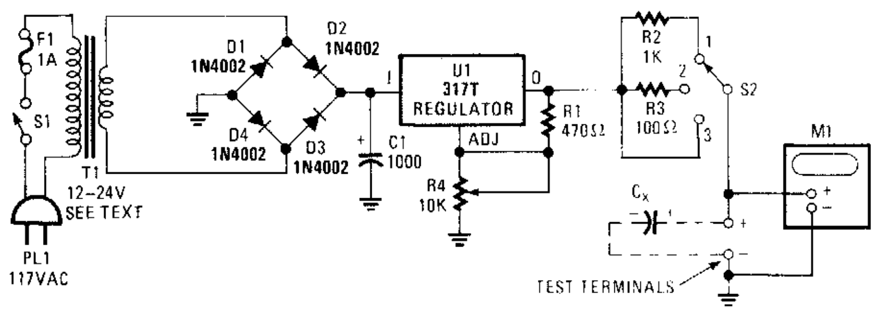


This circuit detects the presence of electrostatic fields by using an FET to alter the flash rate of two LEDs. The FET is installed in the timing circuit and causes a change in the flash rate of the two LEDs.

POPULAR ELECTRONICS

Fig. 46-17

ELECTROLYTIC CAPACITOR REFORMING CIRCUIT

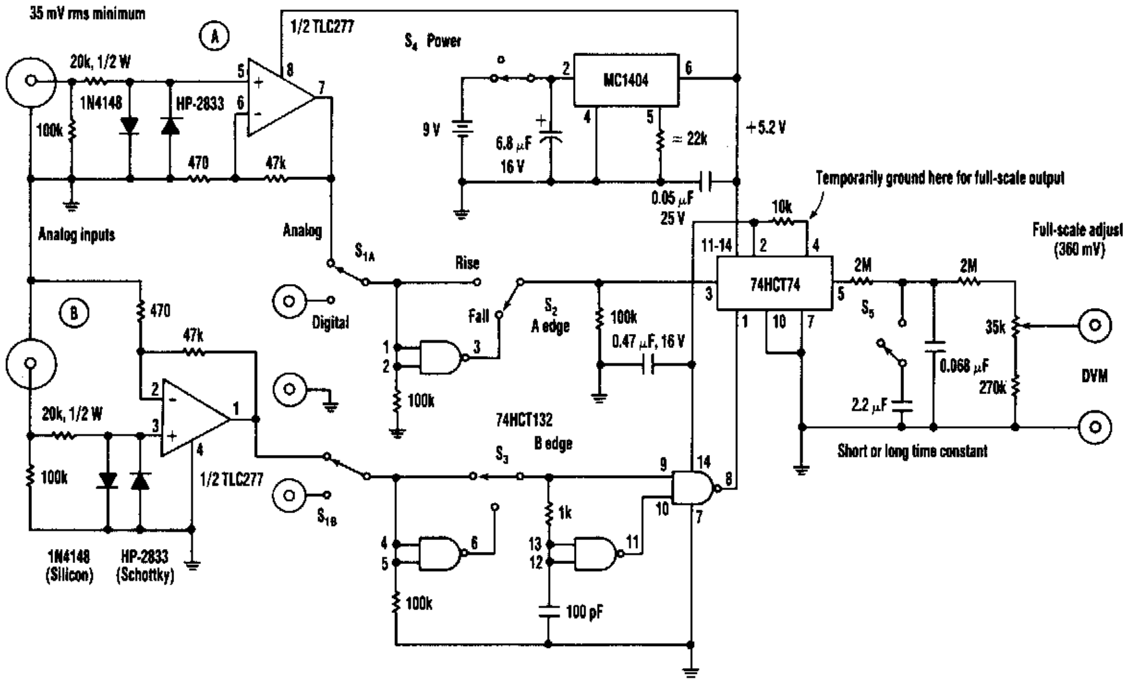


POPULAR ELECTRONICS

Fig. 46-18

Sometimes electrolytic capacitors that are stored for some time will exhibit high leakage currents. Before placing these in service, the capacitors might need to be reformed. This power supply can be used for reforming. Adjust R4 for the capacitor's rated voltage and set S2 in position 1. When M1 indicates the rated voltage, reforming is complete. For large capacitors, use position 2, and then position 3 after reforming starts. T1 can be any transformer with a rating of 12 to 24 Vac at about 1 A.

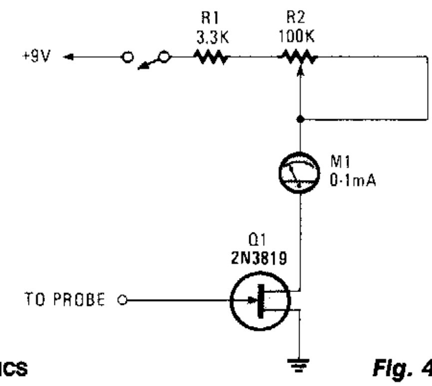
DIGITAL VOM PHASE METER



ELECTRONIC DESIGN ***Fig. 46-19***

The phase-angle meter will work with either analog or digital inputs. A DVM is used as a readout device. The output is 1 mV per degree (360 mV or degrees full-scale). The MC1404 precision regulator maintains calibration with a battery source (9 V).

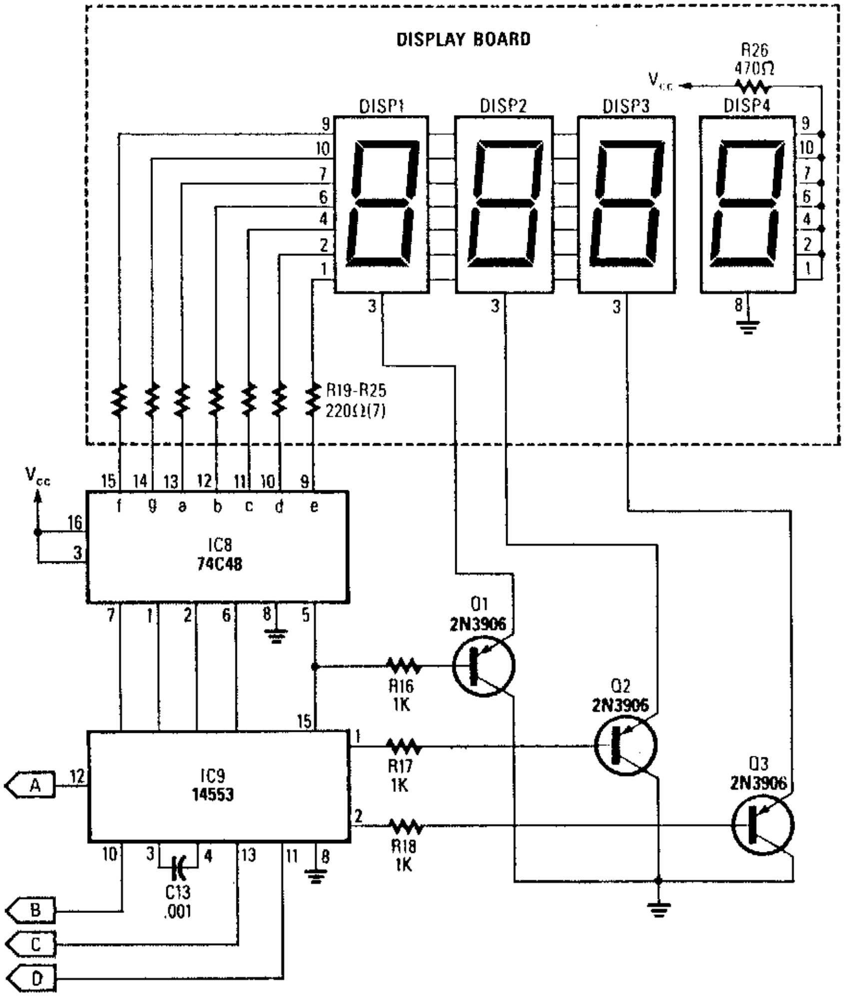
SIMPLE ELECTROMETER



POPULAR ELECTRONICS ***Fig. 46-20***

This electrometer is useful as a relative indicator of static charges or as an electric field in a charge-free environment. An induced negative charge on the probe will reduce drain current toward zero.

DIGITAL TACHOMETER COUNTER

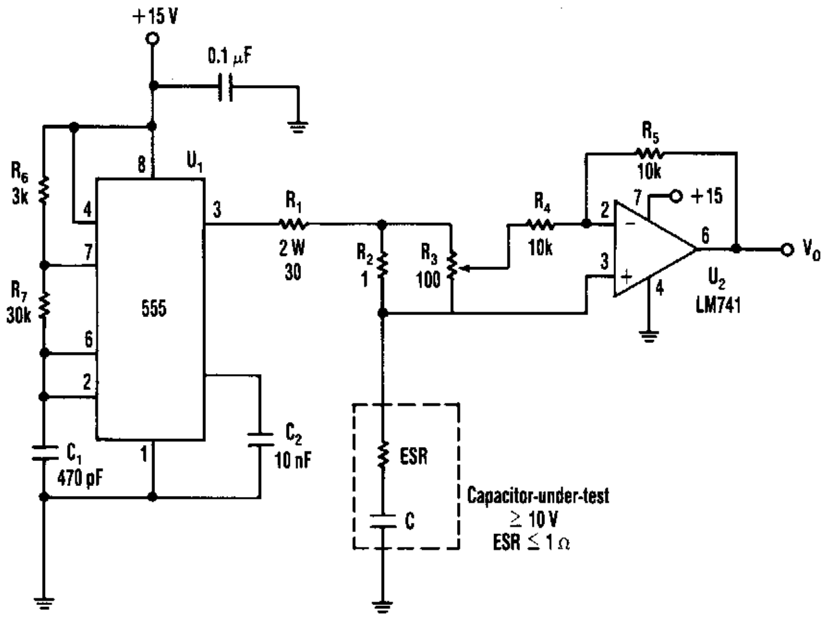


RADIO-ELECTRONICS

Fig. 46-21

This circuit produces a readout for the digital tachometer circuit. IC9 is a 3-digit LED display driver, counter, and latch. IC8 drives the common-cathode LEDs, which are enabled by Q1, Q2, and Q3. See page 268, Fig. 46-5 for the matching project.

CAPACITOR ESR MEASURER



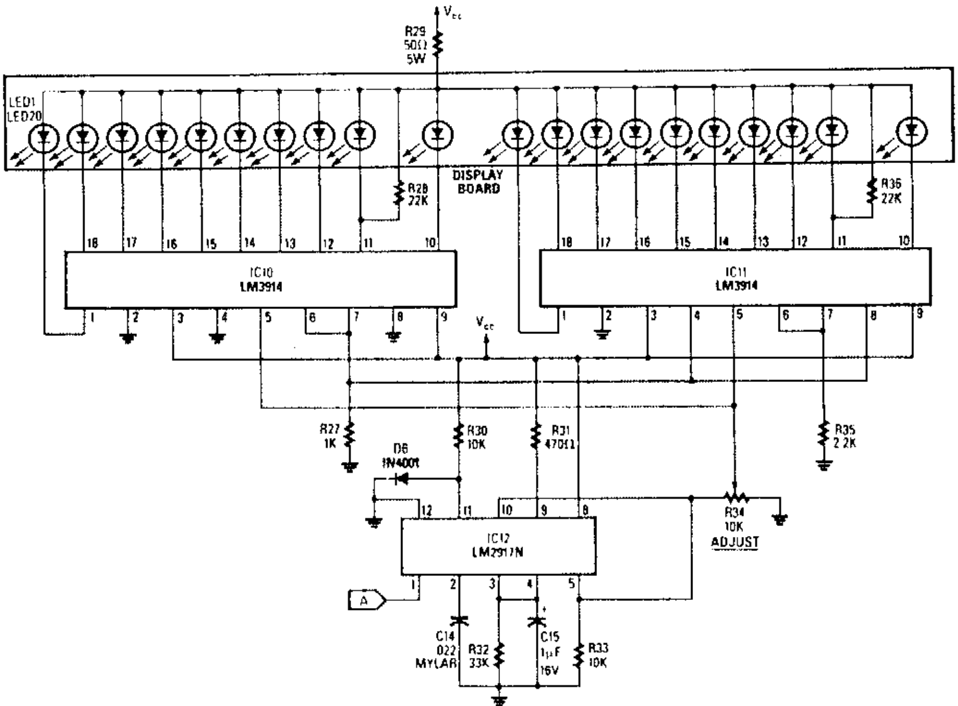
ELECTRONIC DESIGN

Fig. 46-22

The equivalent series resistance (ESR) of a capacitor can be measured using this circuit and an ac voltmeter. U1 functions as a 50-kHz square-wave generator. It drives a current waveform of about ±180 mA in the capacitor-under-test, through R1 and R2. When R3 is adjusted to the proper value, the voltage drop across the equivalent series resistor is precisely nulled by the inverting amplifier (U2). Thus, V_O is the pure capacitor voltage which is the minimum voltage that can be produced at V_O.

To make an ac voltage measurement, adjust R3 until V_O is minimized. Then, note the position of the potentiometer and multiply it by the value of R2, 1 Ω in this case. That product equals the capacitor's ESR. The capacitor is biased at about 7.5 V. Lower-voltage capacitors won't work with this circuit. By changing the value of R2, other ranges of ESR can be measured. However, for small R2 values, the current level should be increased to keep a reasonable voltage across R2. This requires some sort of buffer. The circuit is intended for capacitors greater than 100 μF. The ripple voltage gets large for smaller values and accuracy decreases.

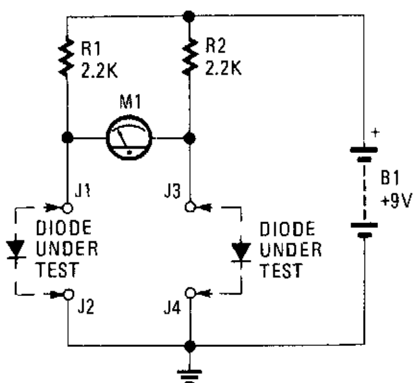
ANALOG TACHOMETER READOUT



RADIO-ELECTRONICS *Fig. 46-23*

The analog display consists of a frequency/voltage converter (IC12) and bar-graph segment drivers IC10 and IC11. R34 is the calibration adjustment and is set so that an engine rpm of 5 000 to 7 000 rpm lights the first LED (redline value).

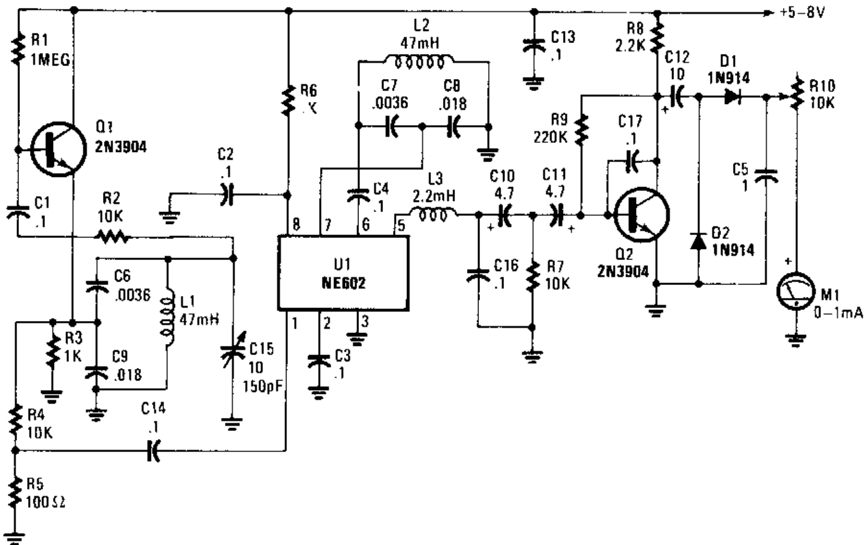
DIODE MATCHING CIRCUIT



This circuit can be used to match diodes for use in circuits where such a balance is necessary (a balanced modulator, for instance). The diode matching circuit will indicate the forward-voltage drop of the two diodes in millivolts.

POPULAR ELECTRONICS *Fig. 46-24*

PERMANENT MAGNET DETECTOR

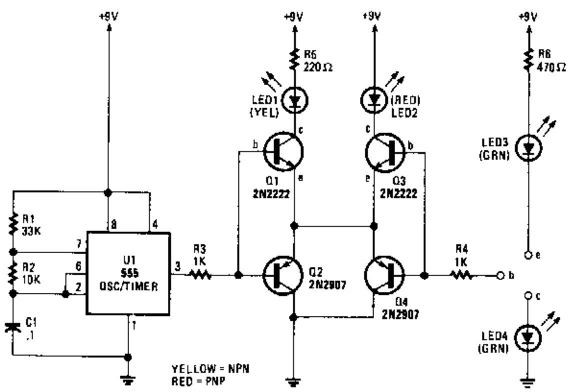


POPULAR ELECTRONICS

Fig. 46-25

In this circuit, oscillator Q1 runs at about 15 kHz and feeds mixer U1. U1 has an internal oscillator that runs at around 15 kHz. C15 is used to zero-beat both oscillators. When a magnet is brought near L1 or L2, the magnetic field shifts the permeability of their respective cases. This changes the oscillator frequency, and the audio note is passed through filter L3, C16, and C10/R7 to amplifier Q2. There the audio is amplified and drives meter M1 via rectifiers D1 and D2.

TRANSISTOR TESTER

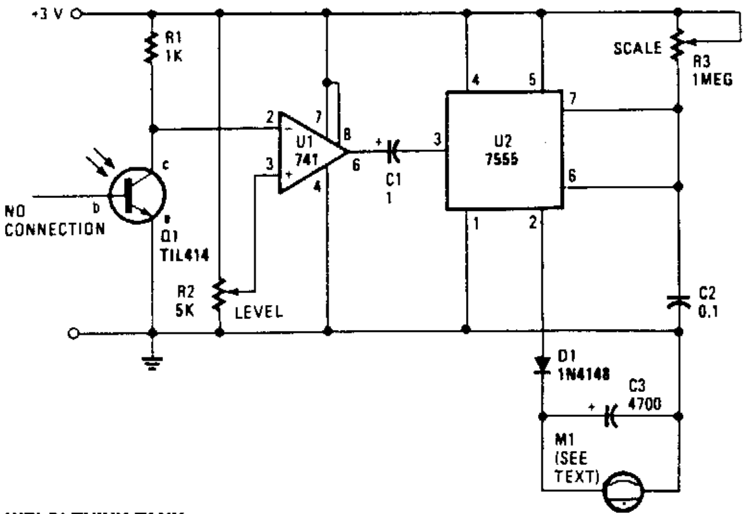


A 555 timer drives Q1 through Q4 with a square wave. LED1 and LED2 light when an npn or pnp transistor respectively, are connected to the text terminals. If LED3 and LED4 light equally as LED1 and LED2, the transistor is functional. If LED3 and LED4 are brighter than LED1 or LED2, the transistor is shorted.

POPULAR ELECTRONICS

Fig. 46-26

BIKE SPEEDOMETER

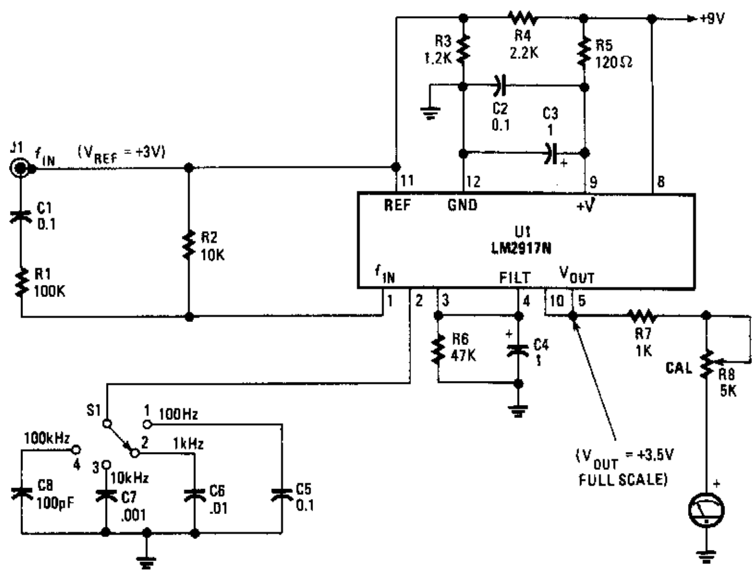


WELS' THINK TANK

Fig. 46-27

A TIL414 photo transistor senses reflection from a spoke-mounted reflector, generates a pulse, and sends it to U1 and U2, a monostable multivibrator, which drives meter M1.

FREQUENCY METER



POPULAR ELECTRONICS

Fig. 46-28

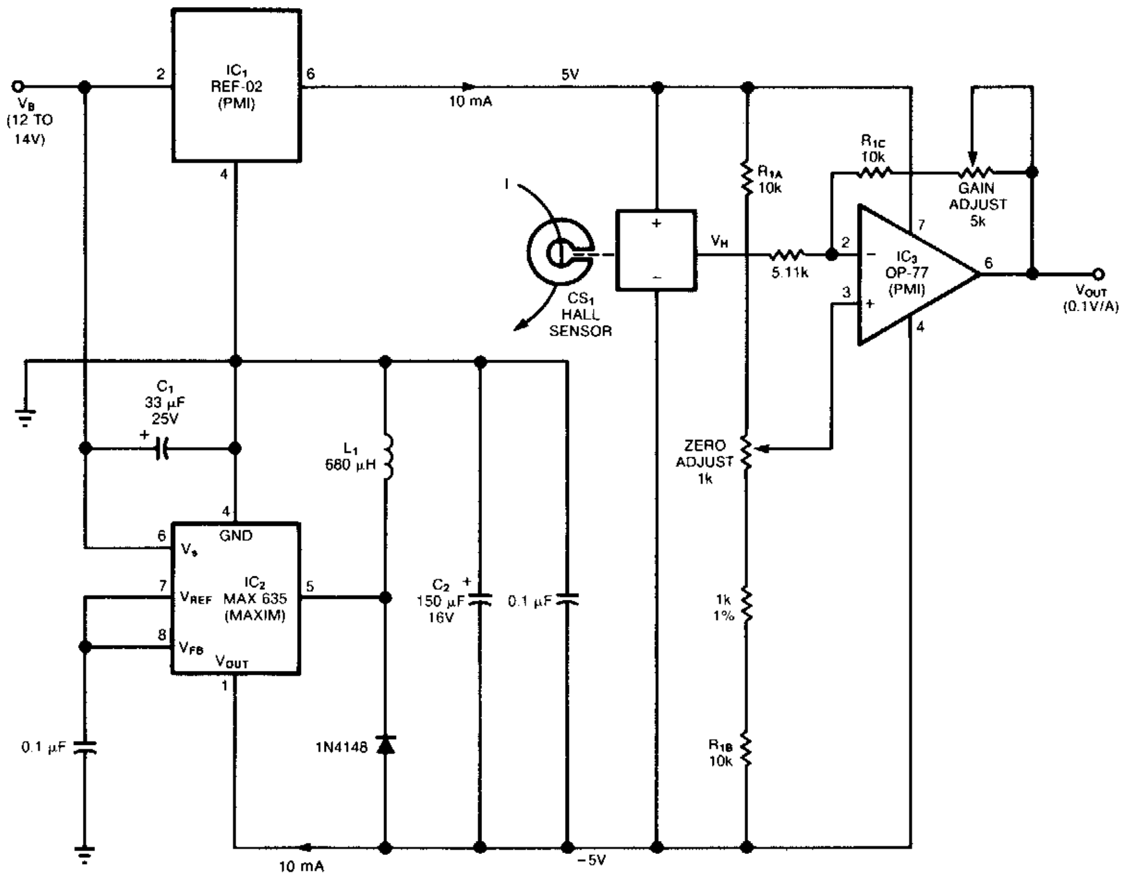
Using an LM2917N, this F/V converter-based circuit indicates the frequency on a meter. S1 selects full-scale range (up to 100 kHz). R8 is recommended to obtain the battery source.

47

Measuring and Test Circuits (*E*, *I*, and *R*)

The sources of the following circuits are contained in the Sources section, which begins on page 669. The figure number in the box of each circuit correlates to the entry in the Sources section.

HALL-SENSOR CURRENT MONITOR



NOTES:

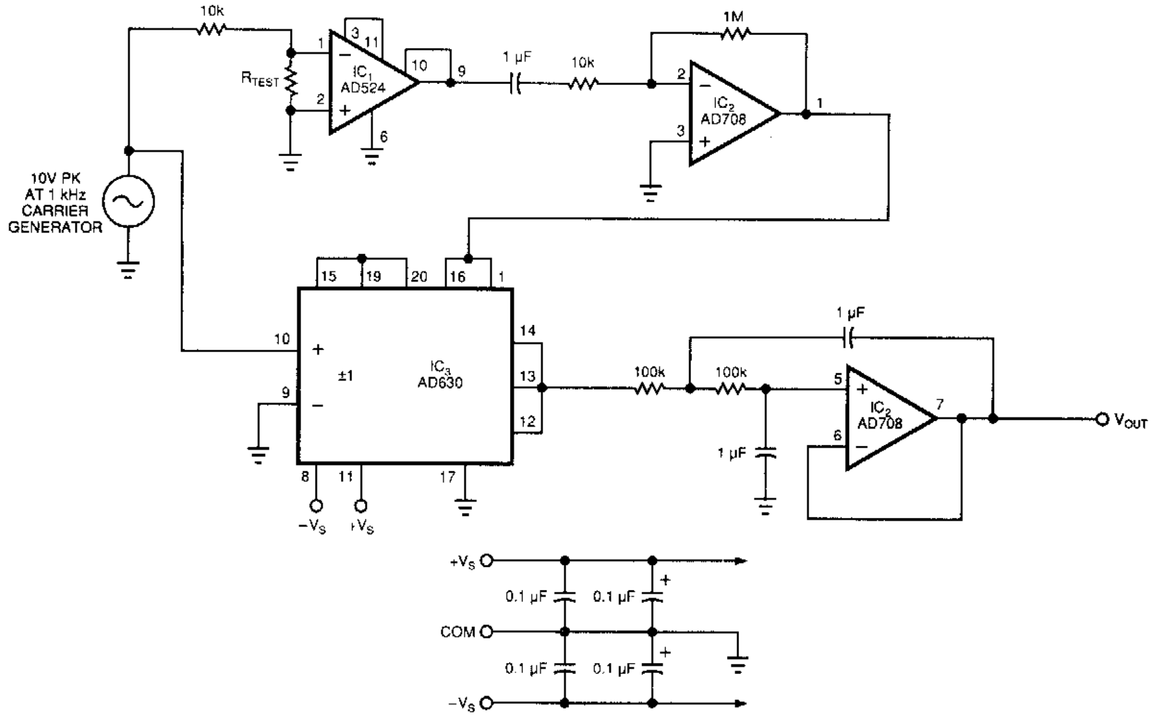
1. C_1 and C_2 ARE 199D TANTALEX CAPACITORS FROM SPRAGUE.
2. L_1 IS A 6860-23 INDUCTOR FROM CADDELL-BURNS.
3. R_{1A}, R_{1B}, AND R_{1C} ARE PART OF A THIN-FILM RESISTOR NETWORK SUCH AS THE CADDOCK T914-10K.
4. CS_1 IS A HALL-EFFECT CURRENT SENSOR (CSLA1CD) FROM MICROSWITCH.

EDN **Fig. 47-1**

The circuit uses a Hall-effect sensor, which consists of an IC that resides in a small gap in a flux-collector toroid, to measure dc current from 1 to 40 A. Wrap the current-carrying wire through the toroid; the Hall voltage V_H is then linearly proportional to the current (I). The current drain from V_B is less than 30 mA.

To monitor an automobile alternator's output current, for example, connect the car's battery between the circuit's V_B terminal and ground, and wrap one turn of wire through the toroid (or, you could wrap 10 turns—if they'd fit—to measure 1 A full-scale). When $I=0$ V, the current sensor's (CS1's) V_H output equals one-half of its 10-V bias voltage, V_H and V_{OUT} are zero when I is zero; you can then adjust the output gain and offset to scale V_{OUT} at 1 V per 10 A.

SYNCHRONOUS SYSTEM



EDN

Fig. 47-2

The circuit uses a synchronous-detection scheme to measure low-level resistances. Other low-resistance-measuring circuits sometimes inject unacceptable large currents into the system-under-test. This circuit synchronously demodulates the voltage drop across the system-under-test and can hence use extremely low currents while measuring resistance.

The 10-V (pk), 1-kHz carrier generator injects a 1-mA reference current into unknown resistor, R_{TEST}. Instrument amplifier IC1 and precision op amp IC2A amplify the voltage across R_{TEST} by a gain of 100 000. Synchronous detector IC3 demodulates this voltage, then op amp IC2B acts a low-pass filter on the demodulated voltage. The low-pass filtering will attenuate all uncorrelated disturbances (such as noise, drift, or offsets), while passing a dc voltage that is proportional to the unknown resistance.

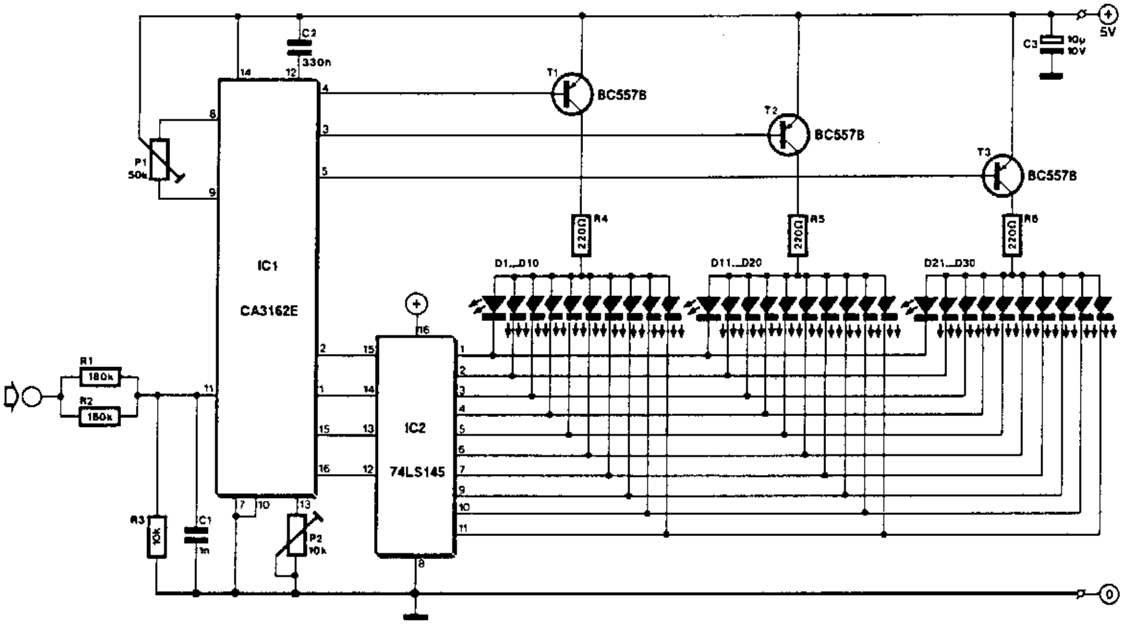
The relationship between the output voltage and the unknown resistance is:

$$V_{OUT} = 10 \times \left(\frac{2\text{ V}}{\pi}\right) \times R_{TEST} \times \frac{10_5}{10\text{ k}\Omega}\text{ ,or}$$

$$R = 0.0157 \times V_{OUT},$$

which is $\frac{15.7\text{ m}\Omega}{\text{V}}$ at the circuit's output.

DIGITAL LED VOLTMETER



ELEKTOR ELECTRONICS *Fig. 47-3*

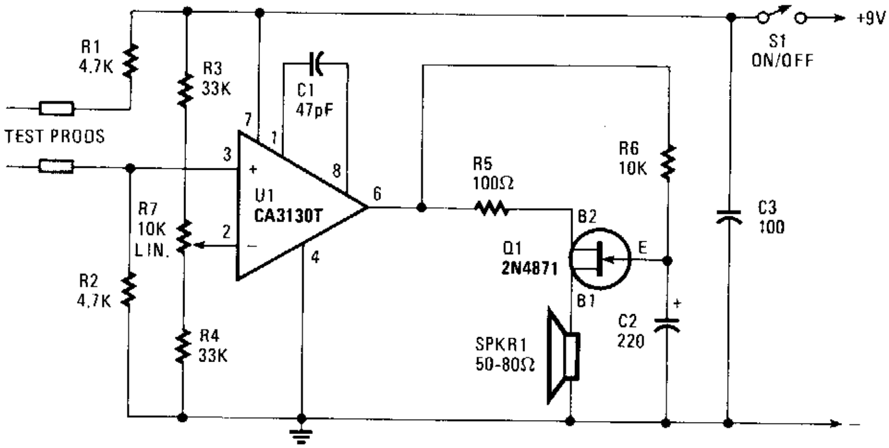
The voltage to be measured is digitized in an analog-to-digital (A/D) converter and then displayed in three decimal digits. The display consists of three groups of 10 LEDs. The meter can only be used for measuring direct voltages.

The A/D converter is based on CA3162, which can process direct voltage up to 999 mV (1-V full-scale deflection—FSD). The FSD is extended to 10 V with the aid of potential divider R1/R2/R3. Other ranges are possible by altering the values of the resistors.

The measured value is read from three bars of LEDs: the first one of these, D1 through D10, shows units; the second, D11 through D20 tens; and the third, D21 through D30, hundreds.

The circuit is nulled with P1 when the input is open. Zero here means that diodes D1, D11, and D21, light. Diodes D10, D20, and D30, represent the figure 9. Next, a known voltage is applied to the input and P2 is adjusted until the LEDs read the correct value. When the input voltage is too high, the display goes out. When the input is negative, the unit LEDs do not light. Notice that variations in the supply voltage affect the measurement adversely.

CONTINUITY CHECKER

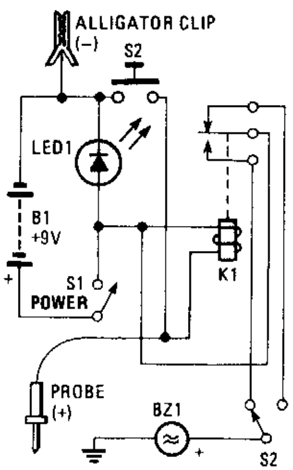


HANDS-ON ELECTRONICS

Fig. 47-4

U1 is an op amp used as a comparator. When the test probes are shorted together, resistors R1 and R2 bias the noninverting input to half the supply voltage. The inverting input is biased by a voltage divider that consists of R3, R7, and R4. Resistor R7 is adjusted so that the voltage to the inverting input is lower than that to the noninverting input when the probes are shorted together. With continuity across the probes, U1's output goes high and supplies power to Q1, which is configured as a relaxation oscillator. The output of Q1 is fed to a high-impedance loudspeaker for an audio tone. When the probe is open, the noninverting input goes to the negative supply rail via R2. This action forces U1's output low, which results in no output from the oscillator.

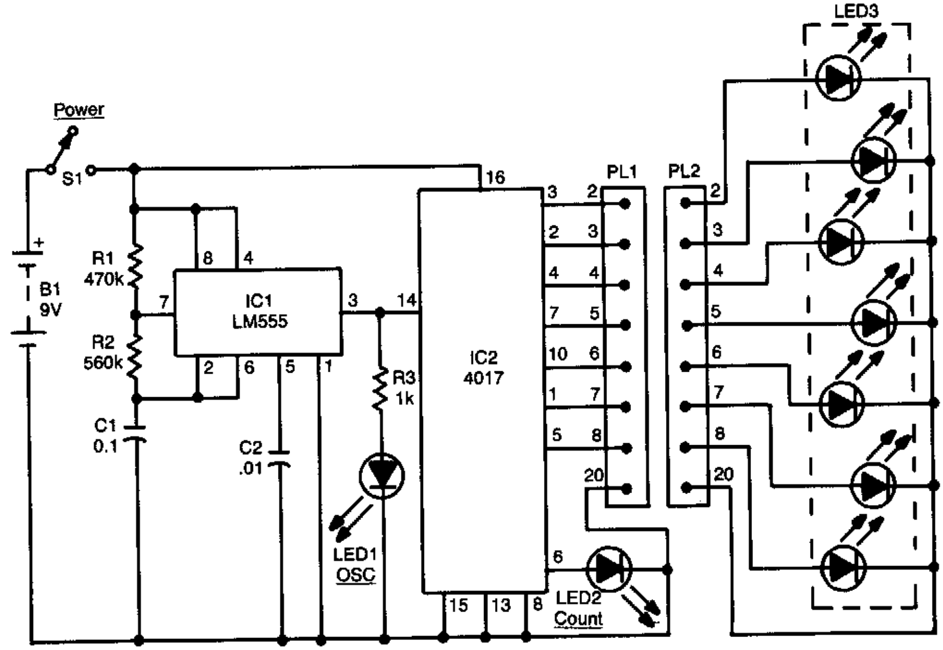
CONTINUITY TESTER



POPULAR ELECTRONICS *Fig. 47-5*

Using an LED and an audible indicator, relay K1 actuates buzzer BZ1. S2 is a buzzer/battery test switch for testing the battery in both NO and NC mode.

QUICK MULTICONDUCTOR CABLE TESTER

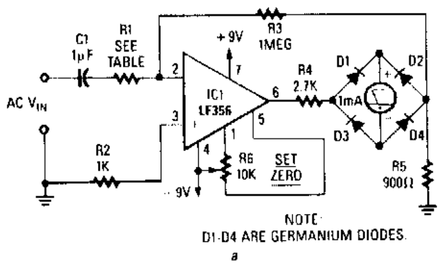


RADIO-ELECTRONICS

Fig. 47-6

This circuit can be used to check up to an eight-conductor cable. IC1, a 555 timer, drives decade counter IC2, a 4017. Each LED should light in sequence. The cable to be tested is connected between PL1 and PL2. If the cable is miswired, the LEDs will light out of sequence. If it is shorted or open, some LEDs will not light.

4-RANGE ac MILLIVOLTMETER



VOLTAGE FULL SCALE	R1 VALUE
1V	1MEG
100mV	100K
10mV	10K
1mV	1K

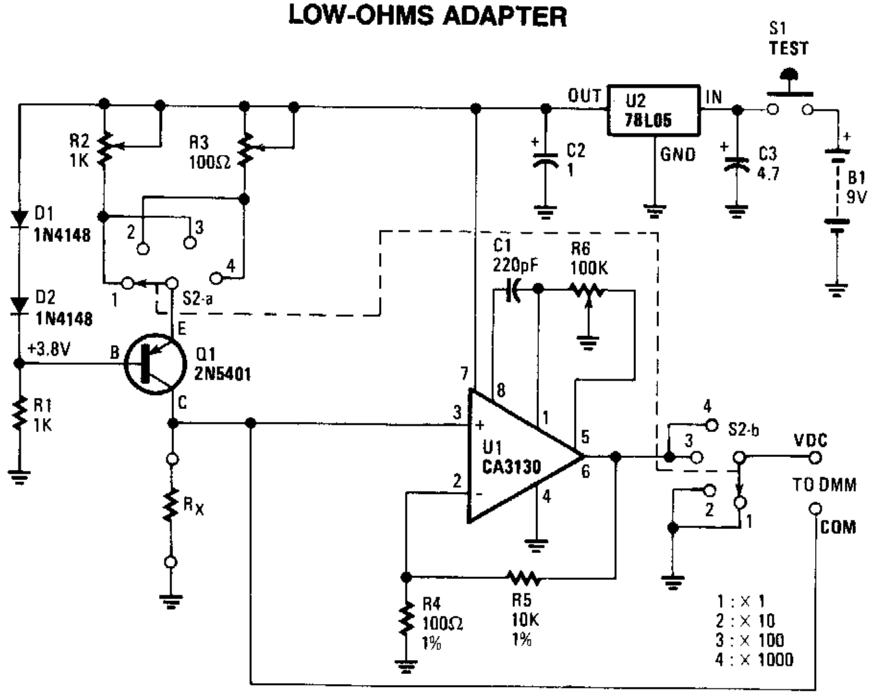
b

RADIO-ELECTRONICS

Fig. 47-7

By placing the rectifier in the op amp feedback path, nonlinearity is greatly reduced to insignificant levels. The meter will read the full-wave rectified average (absolute value) of the input signal. Frequency response is a few Hz to about 50 kHz.

LOW-OHMS ADAPTER



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

Fig. 47-8

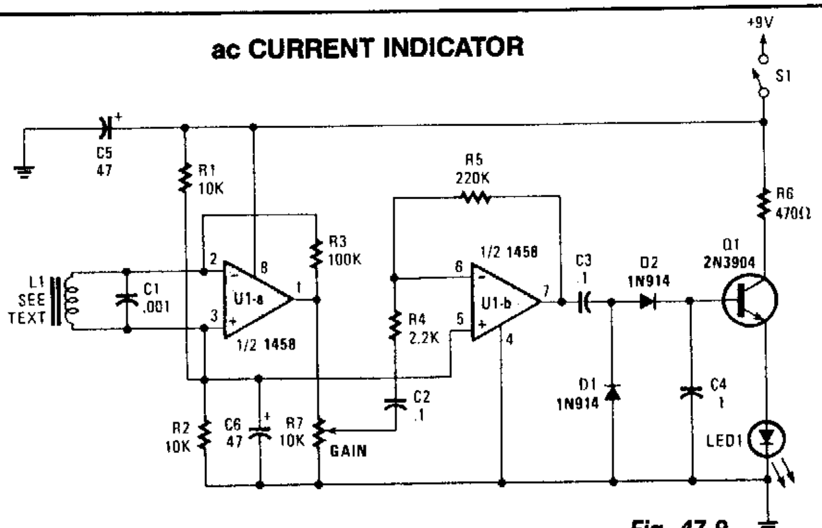
The circuit consists of a 5-V regulator, constant-current source D1, D2, and Q1, and op amp gain stage U1. Power is provided by a 9-V battery whose output is regulated to +5 Vdc by the 3-terminal regulator. The emitter of Q1 is always 0.6 V below the +5-V line. Resistor R1 sets the current through both diodes D1 and D2 to 5 mA.

The resulting 0.6 Vdc across one of the multiturn trimmer potentiometers (R2 and R3), as selected by switch section S2A, sets the current through Q1 and the resistor-under-test.

When R2 is selected, the test current is 1 mA; when R3 is selected, the test current is 10 mA. On the lower two ranges, ×1 and ×10, the voltage across resistance-under-test is applied directly to the DMM terminals. On the upper two ranges, op amp gain stage U1 is switched into the circuit and the DMM measures the voltage between op amp output pin 6 and the test resistor.

When switch S2 is in position 3 (×100) the current set by the constant-current source is 1 mA; the multiplying factor is ×100. When S2 is in position 4, ×1 000, the current is 10 mA and the multiplying factor is 100×10=1 000. Multiturn trimmer-potentiometer R6 adjusts the offset of the op amp so that, with no voltage across the resistor-under-test (i.e., with the measurement terminals short-circuited), the output is zero.

ac CURRENT INDICATOR

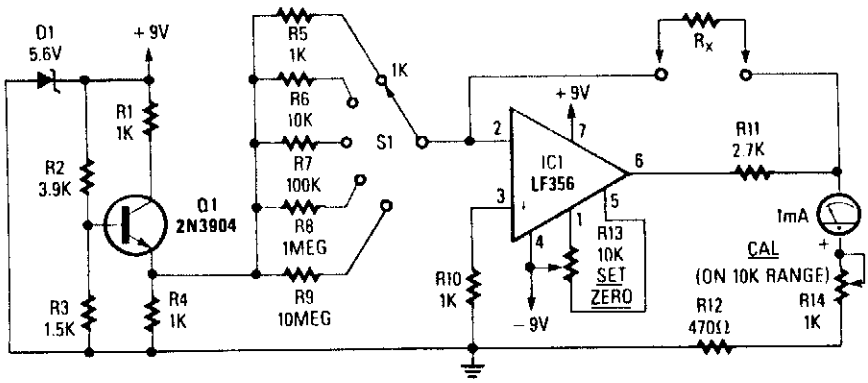


POPULAR ELECTRONICS

Fig. 47-9

Using a dual op amp driving a rectifier and emitter-follower, this circuit indicates ac current on an LED. L1 is an audio transformer winding using a pick-up coil, or 100 to 200 turns of #28 gauge wire 2" in diameter, etc. The circuit can trace ac lines behind walls, etc. or detect ac current flow.

5-RANGE LINEAR-SCALE OHMMETER



RADIO-ELECTRONICS

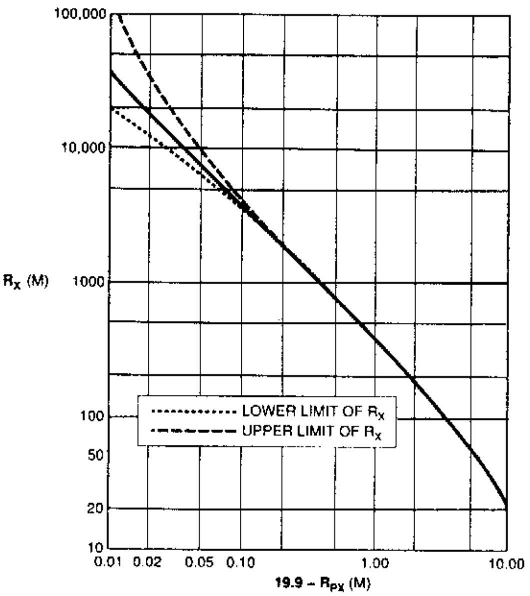
Fig. 47-10

R_x is inserted in the feedback path of IC1. A known reference current is selected from reference voltage generator Q1, D1, and R5 through R9. A meter reading will be produced:

$$I = \frac{R_x}{R_5} \times 1 \text{ mA}$$

where $R_5 = R_5$ through R_9, as selected. This corresponds linearly to the value of R_x.

HIGH-RESISTANCE-MEASURING DMM



EDN

Fig. 47-11

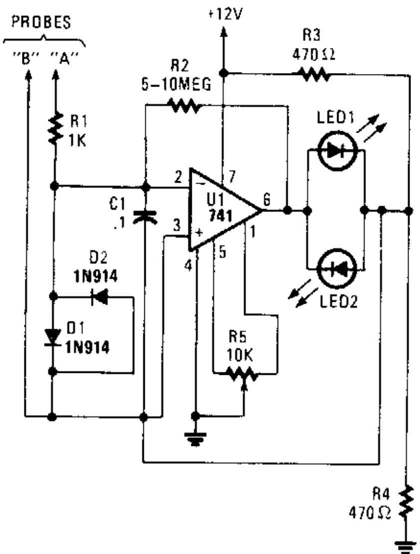
Using a simple technique, you can extend the resistance-measurement range of your 3½-digit DMM from the usual 19.99 MΩ to 40 GΩ. You could measure, for example, the leakage resistances of transformers, motor windings, and capacitors.

For a 19.99-MΩ DMM range, select a stable 20-MΩ resistor whose value is slightly below nominal, 19.99 MΩ for example. An unknown high resistance, R_X, is:

$$R_X = \frac{(R_P \times R_{PX})}{(R_P - R_{PX})}$$

where R_P is the high-value parallel resistor and R_{PX} is the measured value of R_P in parallel with R_X. An even easier way to determine the value of R_X is to use the graph.

HIGH-GAIN CURRENT SENSOR

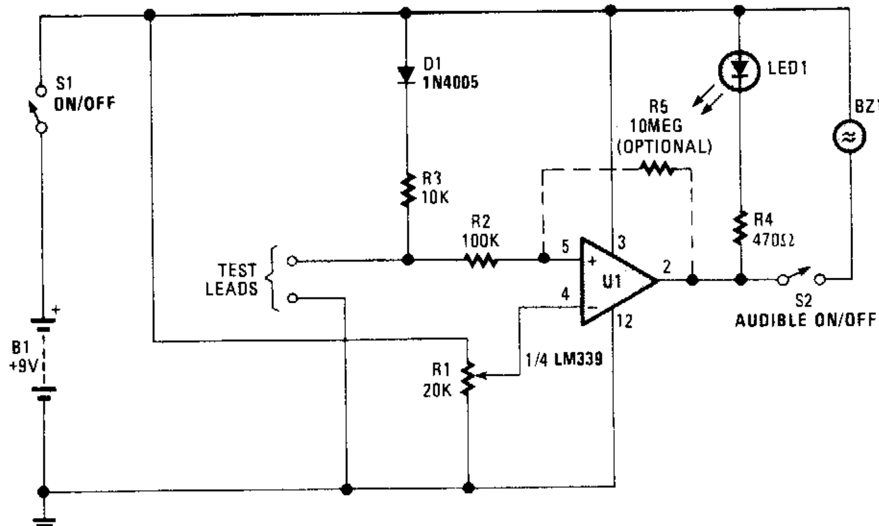


POPULAR ELECTRONICS

Fig. 47-12

A high-gain amplifier using a UA741 is used to sense relative voltage drop in a conductor, and therefore current in the conductor. R_2 can be increased to 10 MΩ for increased sensitivity. LED1 and LED2 provide polarity indication. This circuit can be used to detect current flowing in a PC board trace, and also for locating shorts and opens.

SENSITIVE CONTINUITY CHECKER



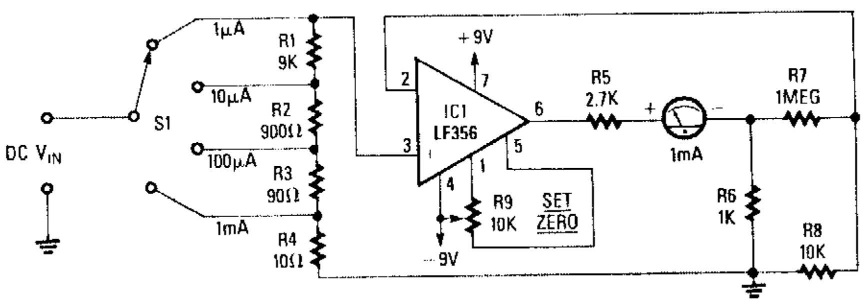
POPULAR ELECTRONICS

Fig. 47-13

This continuity checker (built around an LM339 quad comparator with open-collector outputs) eliminates false readings because of coils or low-resistance devices in a circuit.

U1 is a comparator that acts as a sensing amplifier for the bridge circuit (R1 and D1, R3 and the unknown resistance, R_x that is connected across the test leads. When R_X is less than this predetermined value (by the setting of R1), the LED lights and BZ1 sounds.

4-RANGE dc MICROAMMETER

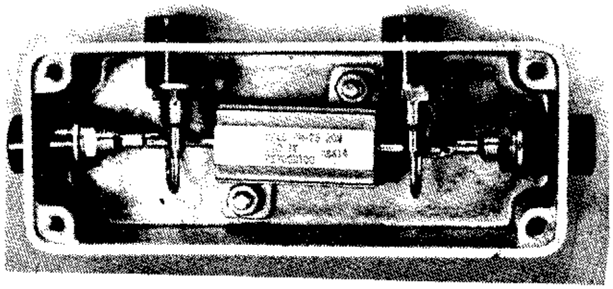


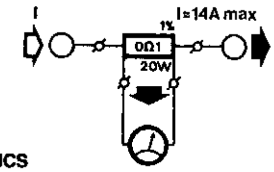
RADIO-ELECTRONICS

Fig. 47-14

IC1 produces a full-scale reading on M1 when pin 3 has a 10 mVdc level on it, as determined by R7 and R8. R1 through R4 are shunts that produce a 10-mV drop for the desired full-scale range. R9 zeros the meter.

MULTIMETER SHUNT



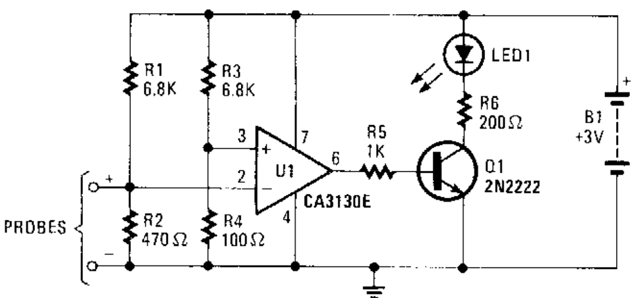


ELEKTOR ELECTRONICS ***Fig. 47-15***

The current range in multimeters, particularly the more inexpensive ones, is restricted by the load limits of the internal shunts to 1 to 2 A. The figure shows how easily a precision heavy-duty resistor from Dale or RCL (0.1 Ω; 20 W; 1%) can be used as an external shunt. These resistors were not designed for this purpose, but they are much cheaper than custom-made shunt resistors. The 20-W rating applies only, by the way, if a heatsink is used: without that its rating is only 8 W.

The maximum current through the device on a heatsink is about 14 A; the larger versions draw up to 17.5 A. When mounting the shunt, make sure that the test terminals (as well as the device terminals) are soldered properly, otherwise the resistance of the terminals is added to the shunt.

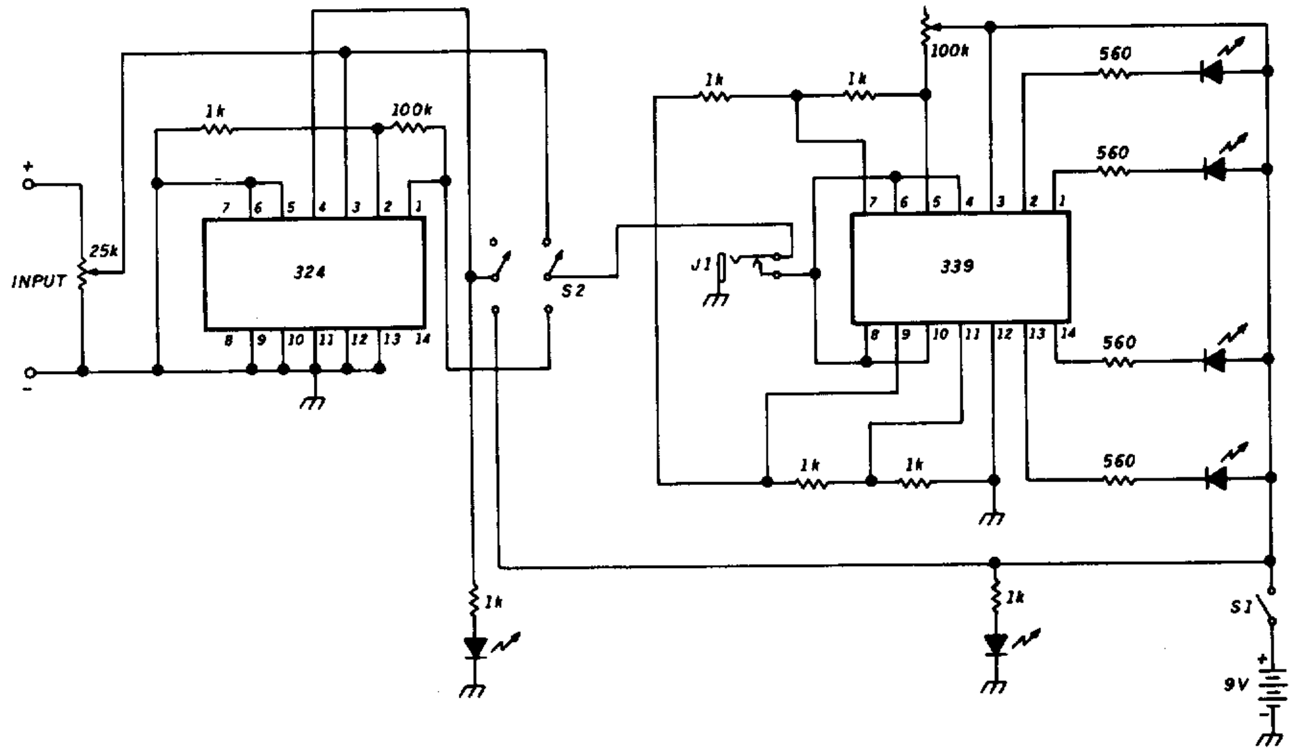
CONTINUITY TESTER



POPULAR ELECTRONICS ***Fig. 47-16***

Using a comparator, this continuity tester applies only about 300 mV to the circuit to be tested. This avoids false readings in semiconductor circuits.

LED MILLIVOLTMETER

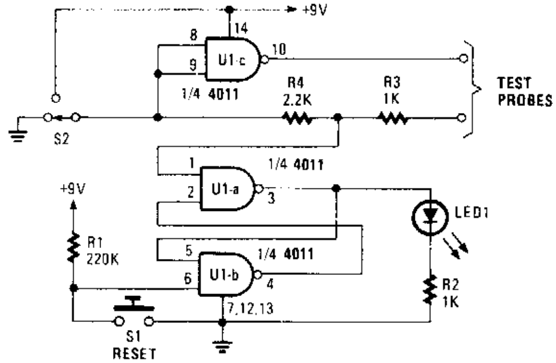


HAM RADIO

Fig. 47-17

This circuit uses a dc amplifier (324) in a bar-graph circuit that uses an LM339 Quad comparator IC to sense dc levels. The LEDs will light every 100 mV. The 324 op amp is configured to provide a gain of 100× to increase this sensitivity to 1 mV. Two auxiliary LEDs indicate "power on" and "×100" gain setting.

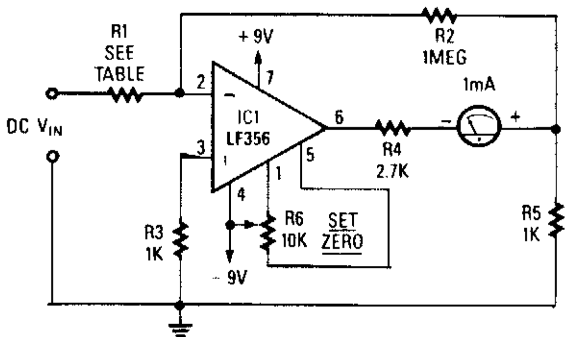
LATCHING CONTINUITY CHECKER



POPULAR ELECTRONICS *Fig. 47-18*

This circuit detects brief shorts or opens. When S2 is in the up position, the circuit indicates if there is or was continuity by lighting LED1. U1A and U1B are connected as an R-S flip-flop. S1 resets the tester. When S2 is in the down position, a momentary interruption in continuity will light LED1. This tester is good for detecting intermittent shorts or opens.

dc MILLIVOLTMETER



a

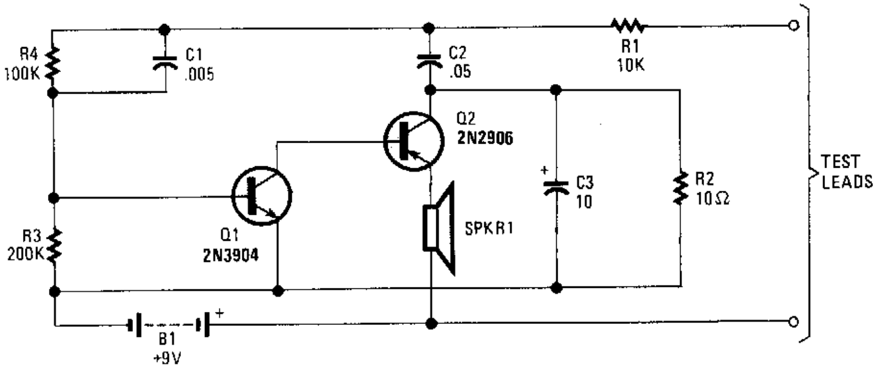
VOLTAGE FULL SCALE	R1
1V	1MEG
100mV	100K
10mV	10K
1mV	1K

b

RADIO-ELECTRONICS *Fig. 47-19*

An LF356 op amp is used as a gain amplifier with the output taken across R5. When a full-scale current of 1 mA is flowing through the meter, exactly 1 V appears across R5 (should be 1% tolerance or better). This is fed back to R2 to the summing junction of IC1 (a full-scale produces 1μA). This offsets the current through R1. R1 has a value of 1 MΩ/V which is used to zero the meter. R4 provides some overcurrent protection for the meter.

CONTINUITY TESTER



POPULAR ELECTRONICS

Fig. 47-20

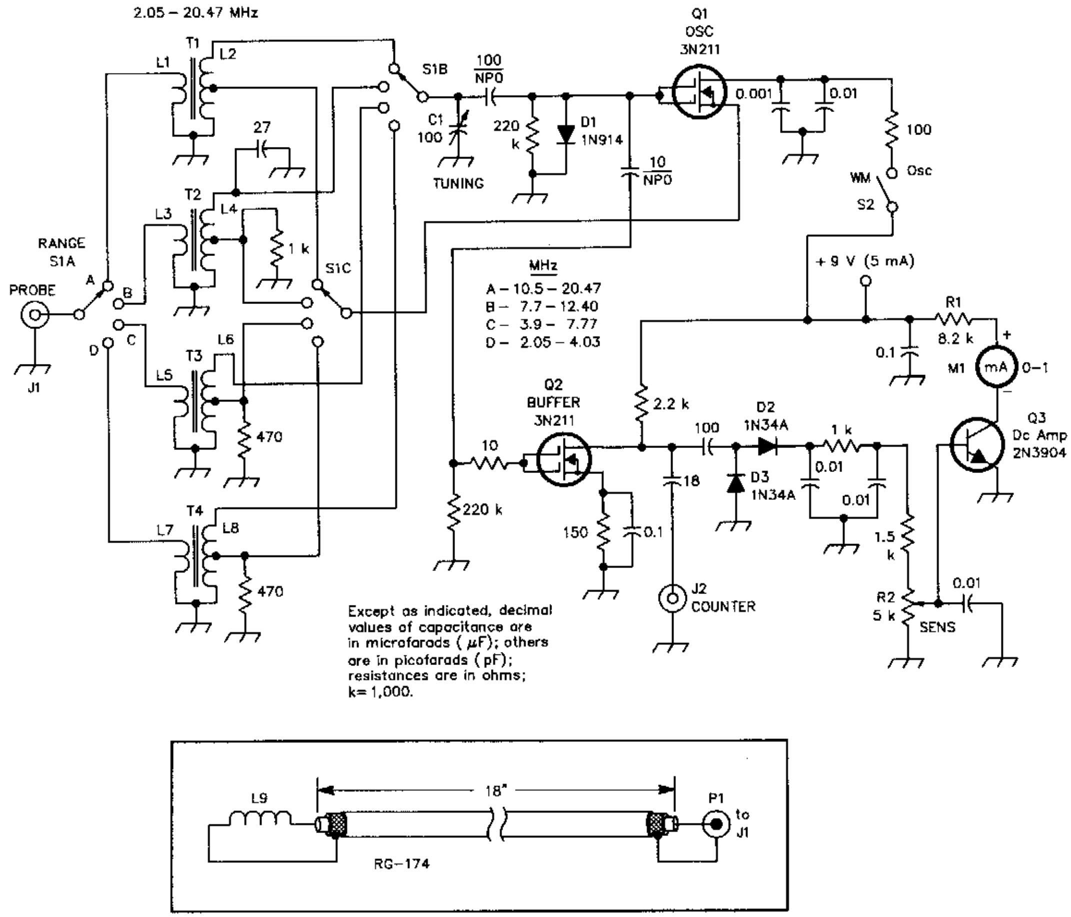
A continuity tester that has an audible indicator can be more useful in some cases than a visual indicator, because you need not take your eyes from the job at hand. Q1 and Q2 form an audio oscillator. When the test leads are connected to a continuous circuit, the oscillator operates, and sounds a tone from the speaker.

48

Measuring and Test Circuits (High-Frequency and RF)

The sources of the following circuits are contained in the Sources section, which begins on page 670. The figure number in the box of each circuit correlates to the entry in the Sources section.

BANDSWITCHED GRID DIP METER

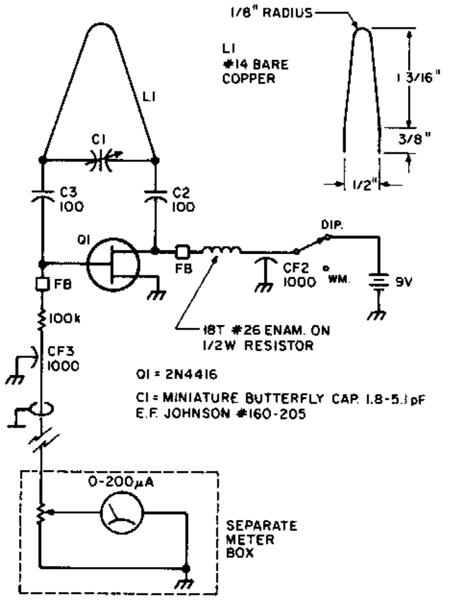


QST

Fig. 48-1

For checking resonances, tuned circuits, antennas, etc., this circuit covers the 2-to 20-MHz range. Q1 serves as an oscillator tunable over this range via C1 and bandswitched coils L1 through L8. When the probe is coupled to a circuit resonant at the oscillation frequency, some RF power will be absorbed and the oscillator output will drop. Q2, D2, D3, and Q3 form an RF detector and dc amplifier to drive meter M1, which will show the drop in Rf level, indicating resonance. R2 is a sensitivity control.

UHF "SOURCE" DIPPER



73 AMATEUR RADIO *Fig. 48-2(a)*

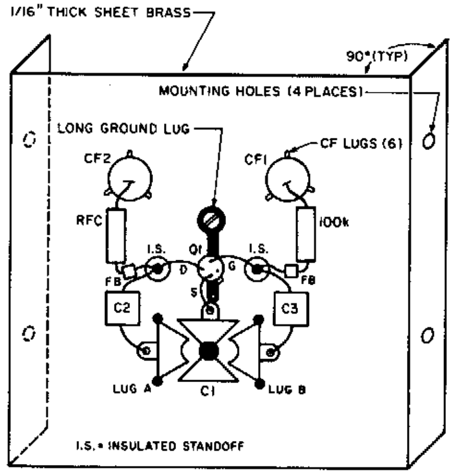
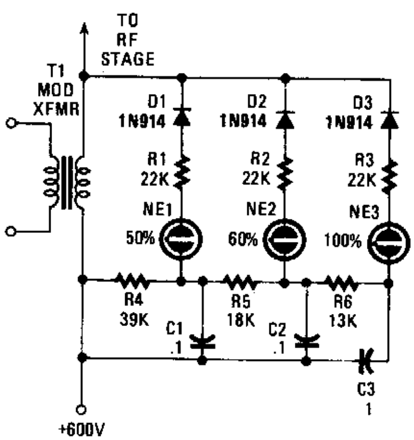


Fig. 48-2(b)

This dipper is useful for UHF experiments in the 420-to-450 MHz amateur band. Because layout can affect performance, follow Fig. 48-2(b).

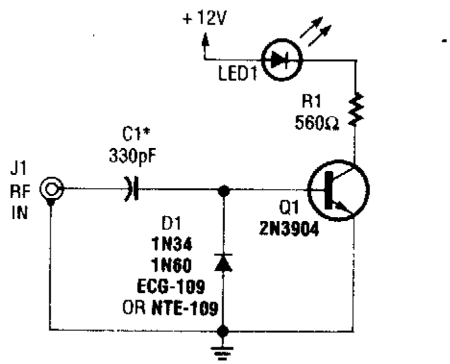
MODULATION MONITOR



POPULAR ELECTRONICS *Fig. 48-3*

Suitable for AM transmitters, this circuit uses neon lamps to indicate 50%, 85%, and 100% modulation on negative peaks.

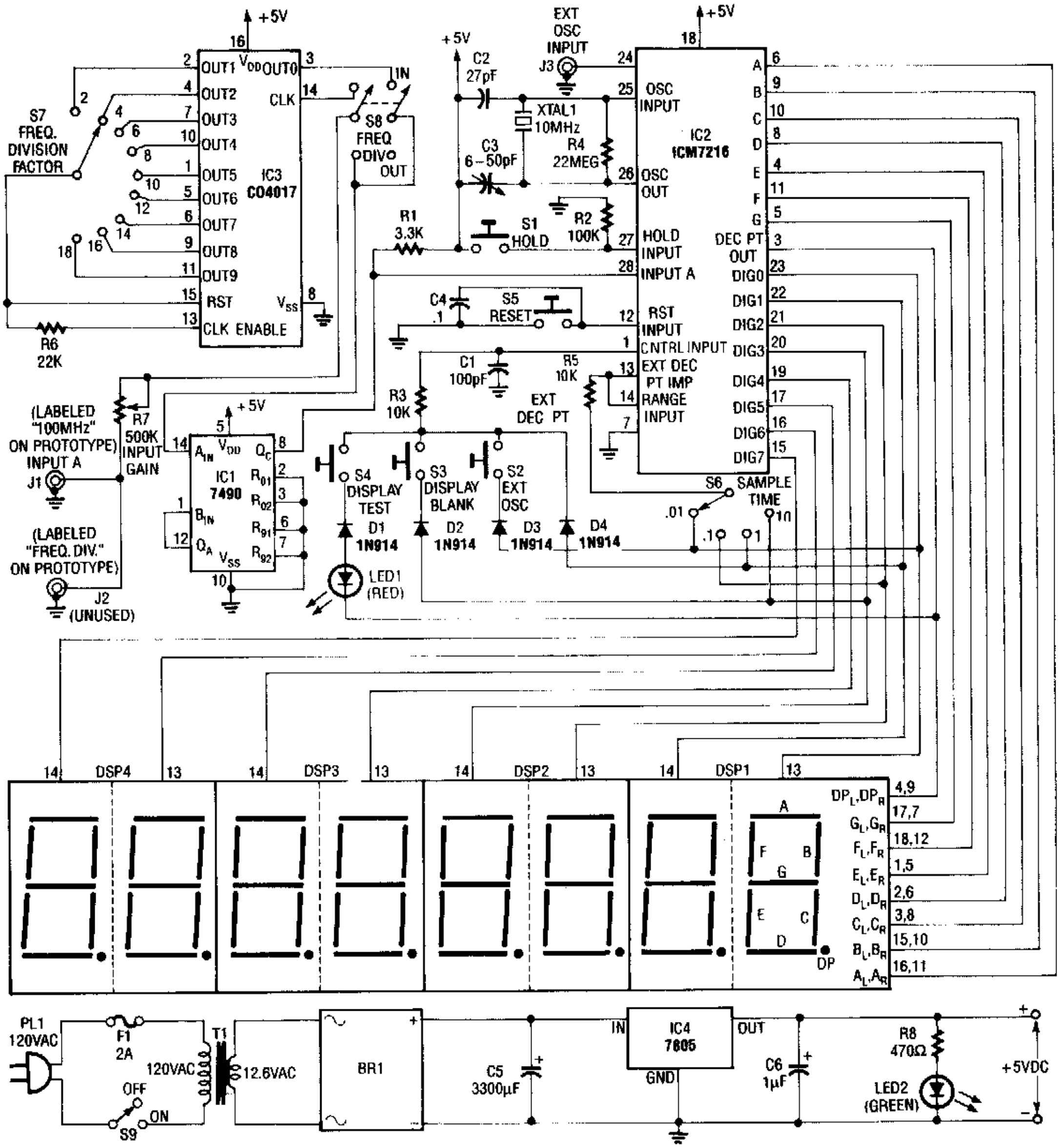
RF OUTPUT INDICATOR



POPULAR ELECTRONICS *Fig. 48-4*

A simple RF detector circuit using a visual indicator can be useful for an RF output indicator, etc. This circuit was used for a transmitter ON indicator.

FREQUENCY COUNTER

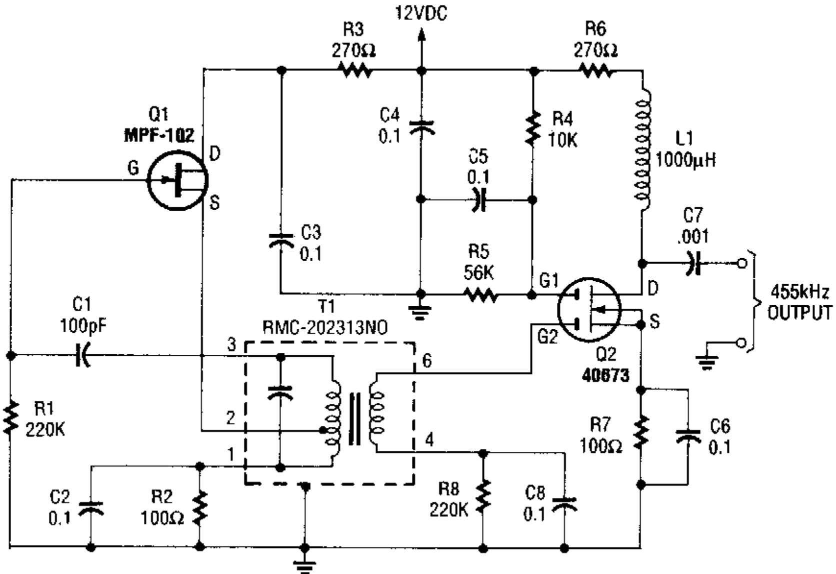


RADIO-ELECTRONICS

Fig. 48-5

Built around an Intersil 7216 frequency-counter IC, this counter has a basic range of 10 MHz, a 100-MHz prescaler, and an extra frequency divider (IC3). This divider divides by an extra factor, as marked on S7 (see schematic), to extend the range of the counter. The display is multiplexed. MAN6710 2-digit red common anode 7-segment LED displays were used on the prototype.

455-kHz AM IF SIGNAL GENERATOR

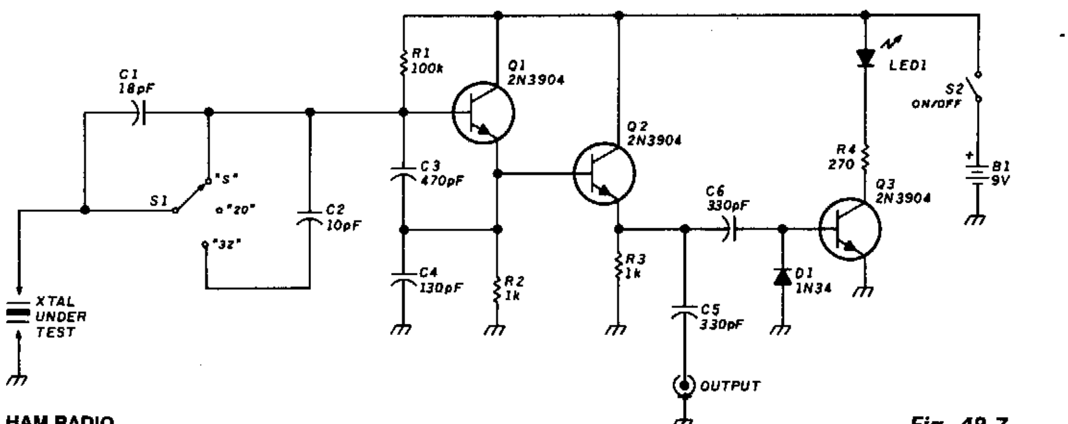


POPULAR ELECTRONICS

Fig. 48-6

An MPF102 FET oscillator drives a dual-gate MOSFET buffer. The MPF102 is configured as a Hartley oscillator. If desired, an audio voltage can be coupled to the junction of R4, R5, and C5 with an extra coupling capacitor ($\approx 1\mu F$) to AM modulate the signal. T1 = Toko P/N RMC—202313NO.

PRECISION CRYSTAL FREQUENCY CHECKER

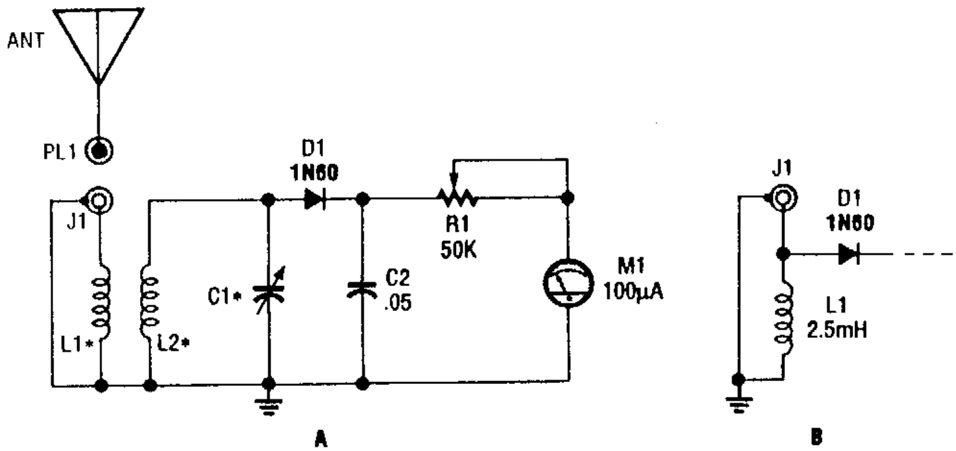


HAM RADIO

Fig. 48-7

This circuit uses a Colpitts oscillator (Q1) with a buffer amplifier (Q2) to test crystals. S1 selects three load conditions—series (S), 20 pF, and 32 pF. Leads to S1 and the crystal should be kept short. The circuit should be useful over 2-to 20-MHz.

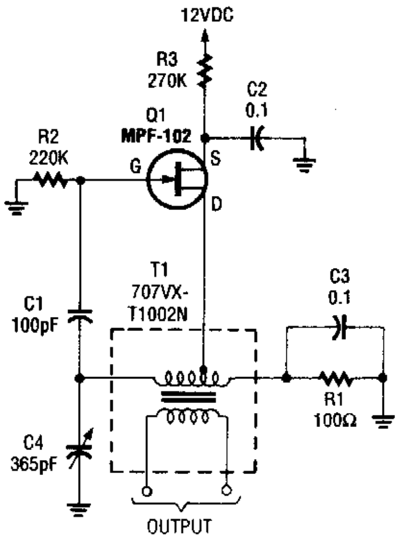
TUNED RF WAVEMETER



POPULAR ELECTRONICS *Fig. 48-8*

L1 and L2 form a tuned transformer. About a 1:3 turns ratio is optimum. L2 and C1 tune to the desired frequency. The frequency range can be 10 kHz to over 200 MHz, depending on the value of C1. For HF use, C1 can be a 140-pF variable. For VHF, use about 25 pF. Use of a 2.5 μH RF choke will yield an untuned wavemeter.

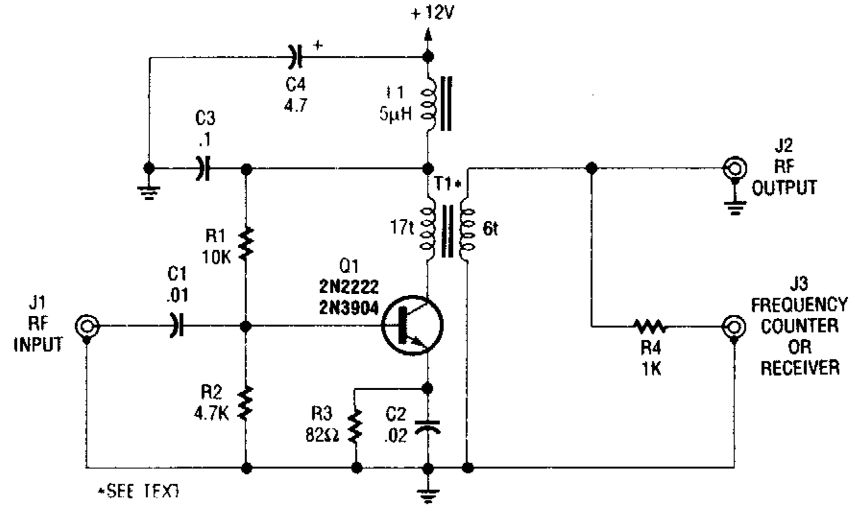
AM BROADCAST-BAND SIGNAL GENERATOR



POPULAR ELECTRONICS *Fig. 48-9*

A Hartley oscillator using an MPF102 covers the band from 530 to 1 600 kHz. T1 is a Toko P/N T1-707VXT1002N 217 μhy transformer.

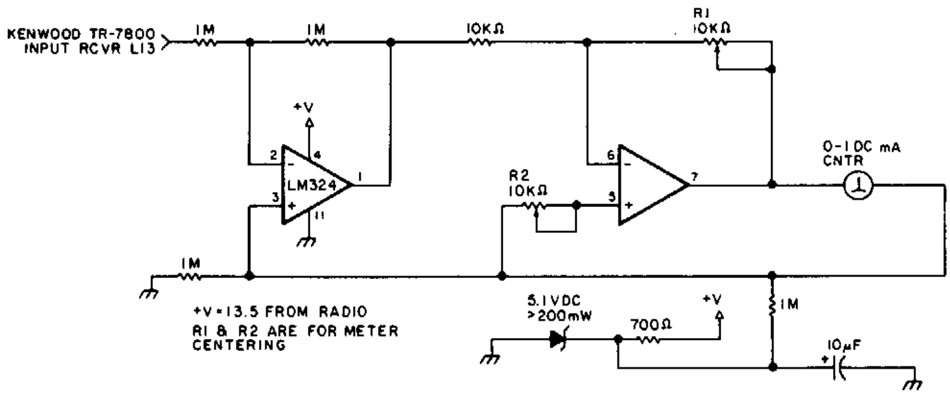
WIDEBAND TEST AMPLIFIER



POPULAR ELECTRONICS **Fig. 48-10**

This single-stage amplifier (using a 2N2222 or 2N3904 general-purpose transistor) is useful for interfacing test instruments. T1 is an Amidon Associates FT-23-43 core wound with 17 and 6 turns of #26 wire. J3 is a lower-level output for a monitoring device (such as a receiver frequency counter or spectrum analyzer).

DEVIATION METER



73 AMATEUR RADIO **Fig. 48-11**

You can use this circuit in most FM VHF receivers; the hookup is off the FM discriminator. Because every signal transmitted has its own deviation signature, this can be a real plus in hunting jammers.

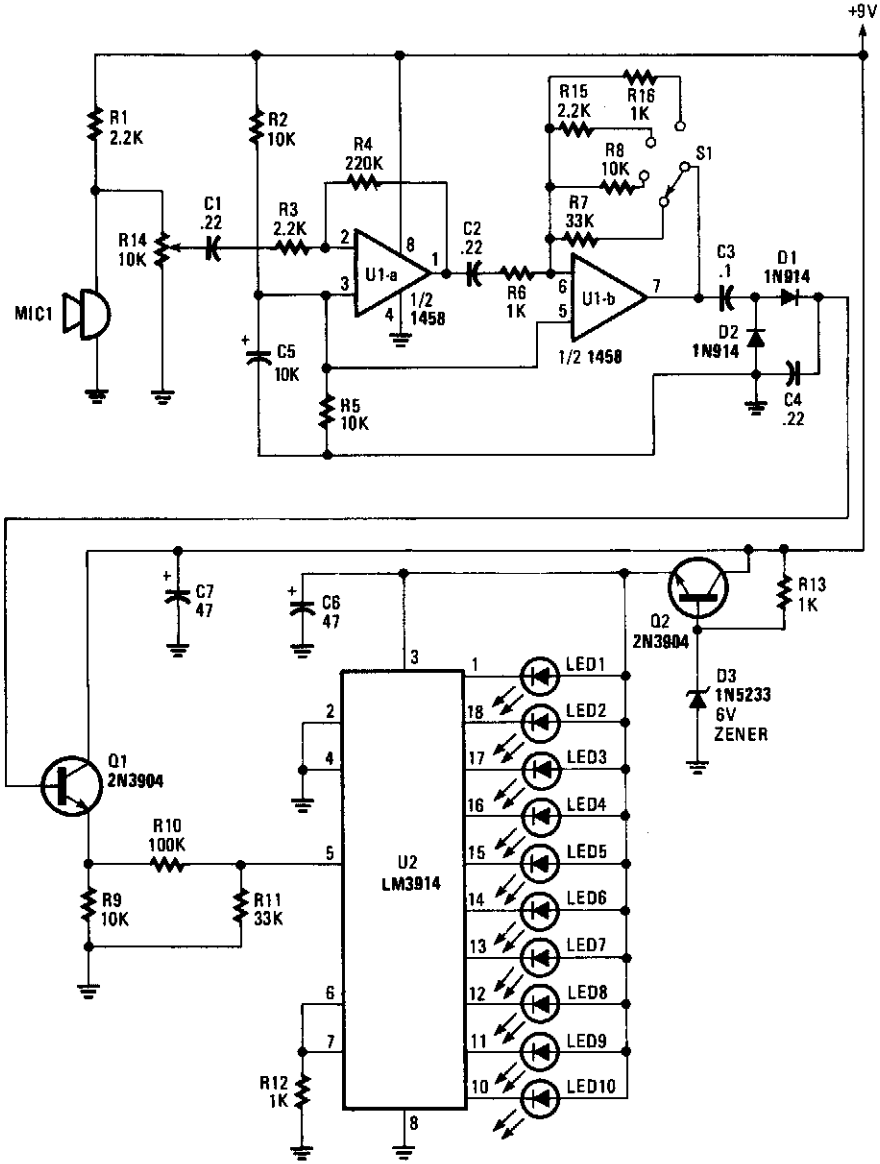
49

Measuring and Test Circuits (Sound)

The sources of the following circuits are contained in the Sources section, which begins on page 670. The figure number in the box of each circuit correlates to the entry in the Sources section.

Sound-Level Meter
Stereo Audio Power Meter
Sound-Level Meter
Noise Generator
Audio Filter Analyzer
Stereo Audio-Level Meter
Mono Audio-Level Meter
Acoustic Sound Transmitter
Acoustic Sound Receiver

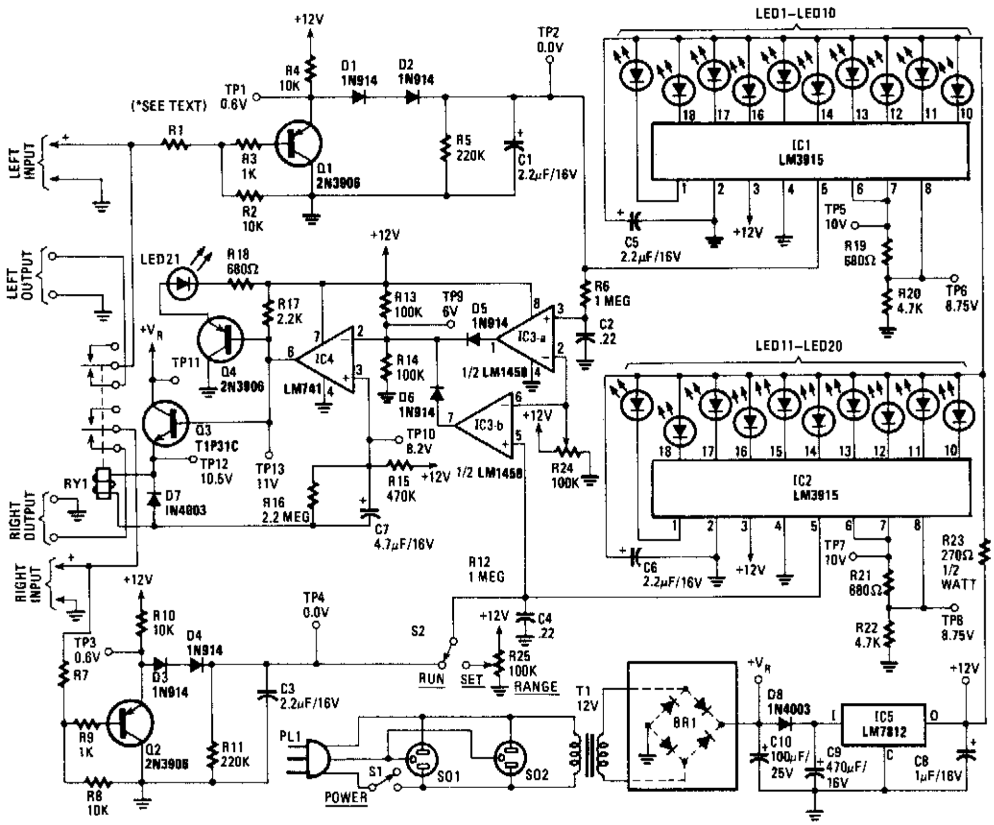
SOUND-LEVEL METER



POPULAR ELECTRONICS *Fig. 49-1*

An electret microphone feeds an audio amplifier/rectifier combination. The amplifier has switchable gain. The rectifier output drives an LM3914 bar-graph generator. R14 provides fine gain control.

STEREO AUDIO POWER METER



RADIO-ELECTRONICS

Fig. 49-2

Attenuation Resistor Values

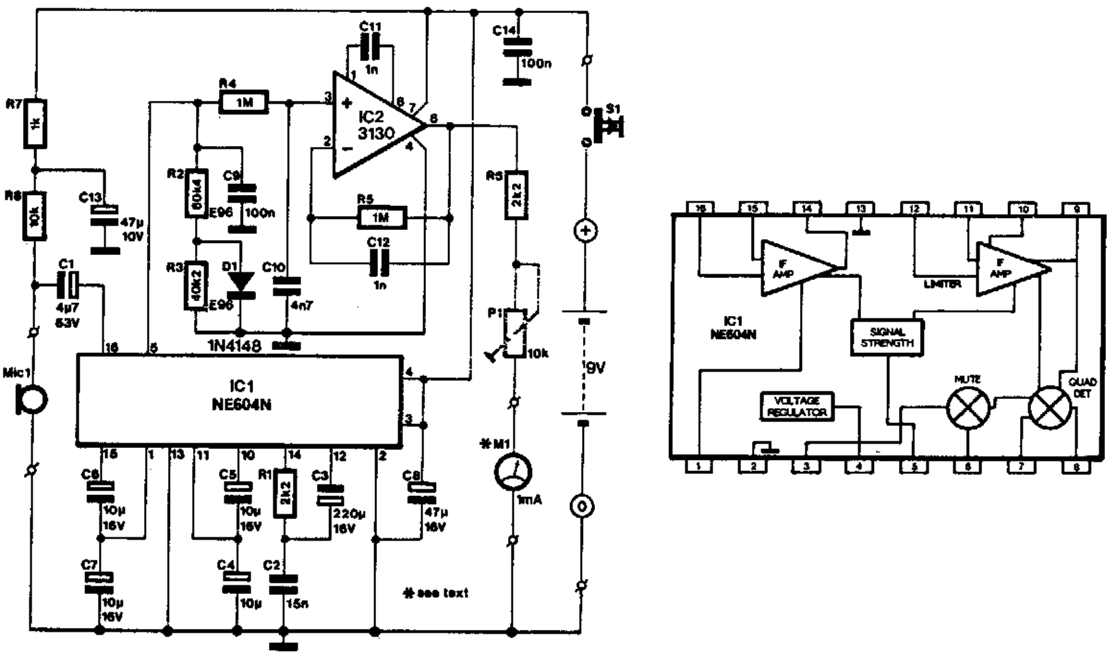
Speaker Impedance	50 W	100 W	200 W	400 W	800 W
2 Ω	*	3.9K	10K	18K	30K
4 Ω	3.9K	10K	18K	30K	47K
8 Ω	10K	18K	30K	47K	68K
16 Ω	18K	30K	47K	68K	100K

(*To read a maximum power level of 50 W into 2 Ω, R1 and R7 should be replaced by a piece of wire between the appropriate printed circuit board pads.)

Peak Power Displayed

LED	50 W	100 W	200 W	400 W	800 W
1,11	0.1	0.2	0.4	0.8	1.5
2,12	0.2	0.4	0.8	1.5	3
3,13	0.4	0.8	1.5	3	6
4,14	0.8	1.5	3	6	13
5,15	1.5	3	6	13	25
6,16	3	6	13	25	50
7,17	6	13	25	50	100
8,18	13	25	50	100	200
9,19	25	50	100	200	400
10,20	50	100	200	400	800

This circuit is used to meter the audio power output of an amplifier feeding a speaker. RY1 is actuated if excess power is fed to the speaker. Two channels are included for stereo applications. R1 and R2 and R3 form an attenuator. When a signal level is reached that produces a voltage across C1, comparator IC3A goes high, and IC4 and Q4 produce enough drive to Q3 to trip relay 1, which cuts off the speakers. LED21 will light as well. In addition, IC1 reads the voltage across C1. IC1 is a bar-graph driver, which lights bar-graph display LED1 through LED10.



ELEKTOR ELECTRONICS

Fig. 49-3

The NE604's signal-strength indicator section is used, based on an internal logarithmic converter. This enables a linear decibel scale so that the moving-coil meter (shown in the diagram) can be replaced by a digital instrument.

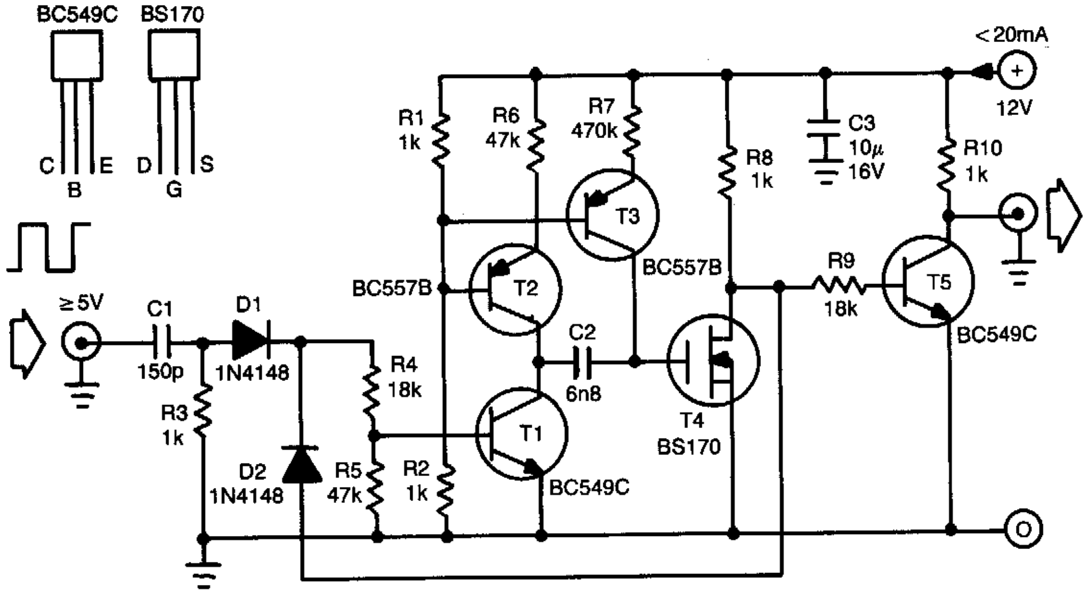
The signal source is assumed to be an electret microphone that converts ambient noise into an electrical signal. Because this type of microphone normally contains a buffer stage, R7, R8, and C13 have been included to provide the supply voltage for this stage.

The NE604 delivers an output current (at pin 5) of 0 to 50 μA, which causes a potential difference across R2 + R3 of 0 to 5 V. The input and output signal range is equivalent to a sound range of 70 dB. To compensate for the effects of temperature changes, the required resistance of 100 kΩ is formed by two resistors (R2 and R3) and a diode (D1).

Any ripple remaining on the output voltage is removed by R4/C9/C10 before the output is buffered by IC2. The indicating instrument, here a moving-coil meter, is connected to the output (pin 6) of IC2 via a series resistance, $R_6 + P_1$. The preset is adjusted to give full-scale deflection (FSD) for an output voltage of 4 V.

Calibrating the meter is a little tricky, unless you have access to an already calibrated instrument. Otherwise, if you know the efficiency of your loudspeaker, that is, how many decibels for 1 W at 1 m, you can use that as reference. The scale of the meter can then be marked with the (approximate) value. In any case, the meter deflection must at all times be seen as an indication, not as an absolute value: it was not thought to be worthwhile to add a filter to the circuit to enable absolute measurements to be made.

NOISE GENERATOR



ELEKTOR ELECTRONICS

Fig. 49-4

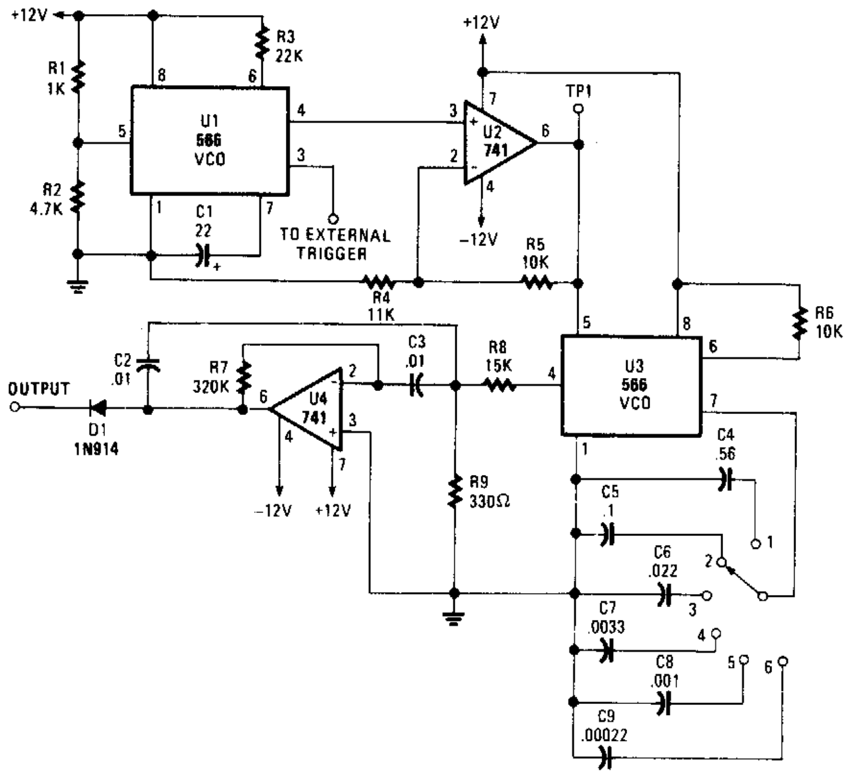
This noise generator provides constant noise energy over its bandwidth, which results from the nonlinear behavior of its switching components, more particularly T4. It is very useful for measurements where limited noise bands are required. Varying the ratio R_6:R_7 and the clock frequency enables the generated noise to be adapted to specific requirements.

Transistors T2 and T3 are current sources. The current through T2 is about 10 times the level of that through T3. Assuming that T4 is on and that the clock input is low, T1 is off, and C2 discharges. The capacitor is pulled to about half of the supply voltage by the two current sources. When that state is reached, stability ensues because the potential then present at the gate of T4 keeps the FET switched on.

When the clock goes high, T1 is switched on so that C2 is connected between the gate of T4 and the earth. Because C2 is only partly charged, the FET is switched off. Transistor T1 is kept switched on by OR gate D1/D2 so that the clock pulses are blocked. Capacitor C2 then charges via T3 until the potential across it becomes high enough to switch on T4. Transistor T1 is then switched off and the circuit is ready to receive another clock pulse (or rather a leading edge of one).

Because it is not known when the clock pulse arrives, it is not known to what potential C2 will be discharged by T2 (and countered by T3). It is therefore also not known when the next clock pulse will arrive. In other words, the pulse width of the output signal is varying constantly, which is characteristic of a noise signal.

AUDIO FILTER ANALYZER



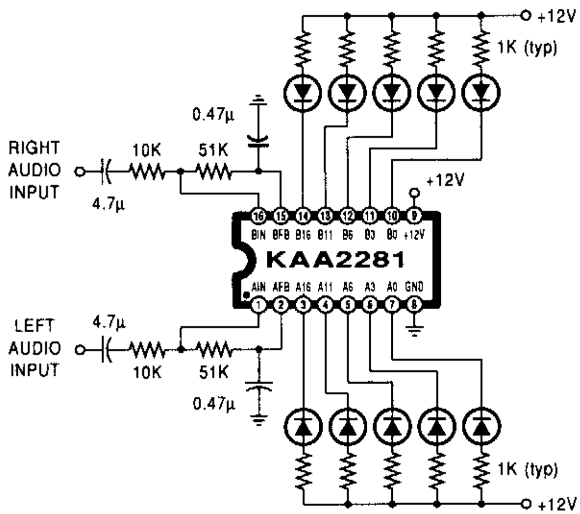
POPULAR ELECTRONICS *Fig. 49-5*

When this circuit is connected to a filter and an oscilloscope, the scope displays the filter's frequency response. A frequency that sweeps from low to high is applied to a filter. An oscilloscope is triggered by the start of the sweep and ends its trace at the highest frequency of the sweep. The filter output goes to the vertical amplifier of the oscilloscope. Using bandpass filters as an example, as the bandpass frequency is approached, reached, and passed, the scope follows the peaking output and draws the response curve. A neat effect!

The 566 VCO (U1) produces a VLF triangle wave to frequency modulate the next stage. It also produces a square wave to externally trigger the scope. Op amp U2 (a 741 unit) optimizes the amplitude and the dc component. Another VCO (U3) produces the actual sweeping triangle wave. Its frequency is selectable via S1. Op amp U4 (another 741 op amp) is set up as a bandpass filter and has been included as an example filter. Finally, diode D1 chops off the bottom half of the output, and leaves a nice bell curve.

To set up and operate, power-up the circuit and scope. Set the scope's TIME/CM to 50 ms/cm. Set the VOLTS/CM control to 2 V. Attach a probe from the circuit's trigger to the scope's external trigger input. Set the triggering mode to normal, external. Attach a probe from the vertical amplifier to TP1. You'll see a diagonal line that runs across the CRT. Input coupling should be set to dc. Adjust the triggering level until the diagonal runs from the upper left to the lower right of the CRT to ensure a displayed sweep from low to high. Now, disconnect the probe from TP1 and attach it to the filter output past the diode.

STEREO AUDIO-LEVEL METER

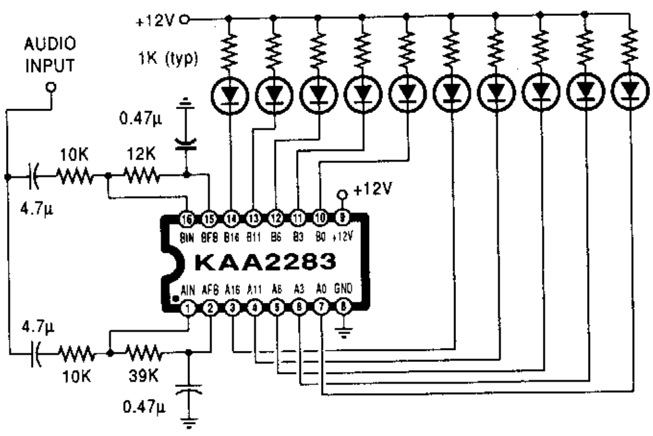


RADIO-ELECTRONICS

Fig. 49-6

A Samsung KAA2281 and a few LEDs make up a simple stereo indicator. Levels displayed are −16, −11, −6, −3, and 0 dB. Input sensitivity is 1 mV. LEDs can be any suitable types or a bar-graph display.

MONO AUDIO-LEVEL METER

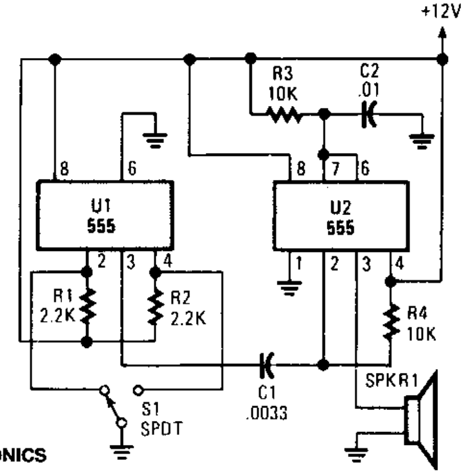


RADIO-ELECTRONICS

Fig. 49-7

This mono indicator uses both halves of a Samsung KAA2283. Levels displayed are −18 to 0 dB in 2-dB steps. Sensitivity is 0.1 to 0.9 mV.

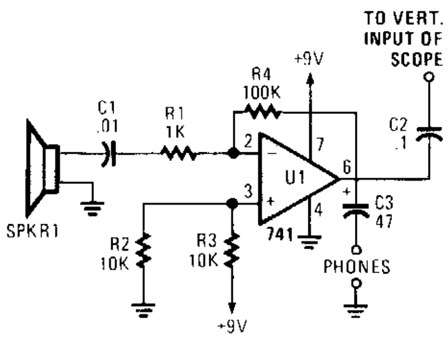
ACOUSTIC SOUND TRANSMITTER



POPULAR ELECTRONICS *Fig. 49-8*

Pulsed sound is produced by this circuit. U1 is used as a bistable multivibrator, which acts as a contact "debouncer" for S1. C1 feeds a trigger pulse to U2, which feeds a pulse to SPKR1, to piezo transducer. Values are shown for a pulse width of 110 μs.

ACOUSTIC SOUND RECEIVER



POPULAR ELECTRONICS *Fig. 49-9*

The receiver is an audio amplifier fed by SPKR1, a piezo speaker that is used as a microphone. A scope or headphones can be used as a detector. The scope can be triggered horizontally by the transmitted acoustic pulse; the vertical display can be used to drive the delay time, and hence the distance.

$$\text{distance (feet)} = \frac{\text{delay time (ms)}}{1100}$$

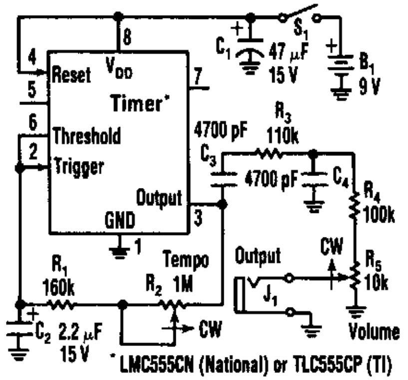
(at 25°C air temperature)

50

Metronomes

The sources of the following circuits are contained in the Sources section, which begins on page 670. The figure number in the box of each circuit correlates to the entry in the Sources section.

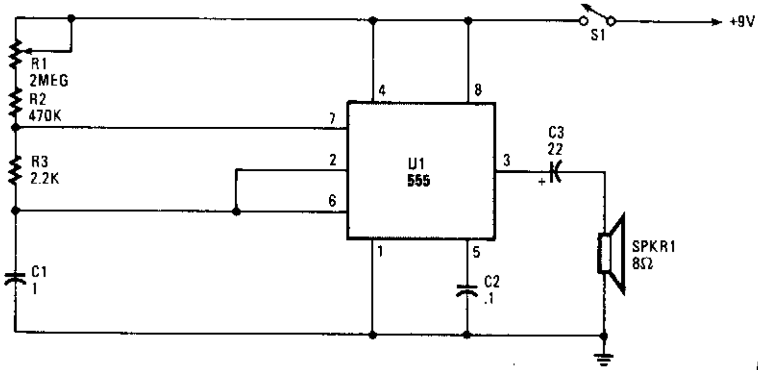
LOW-POWER METRONOME



ELECTRONIC DESIGN Fig. 50-1

Using only 0.25 mA, this metronome is ideal for battery operation. The tempo range is 34 to 246 beats per minute. Use a CMOS timer, such as an LM555 CN or TLC55CP.

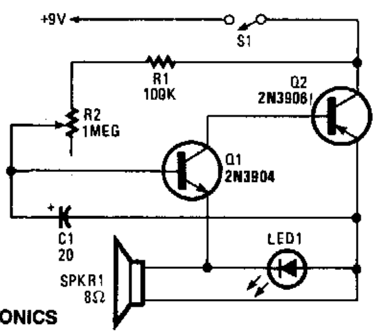
METRONOME I



POPULAR ELECTRONICS Fig. 50-2 .

This simple metronome circuit offers a range of speeds from *largo* to *prestissimo*! The parts values are set so that the repetition rate adjusted by R1 runs from nearly 45 to 184 per minute.

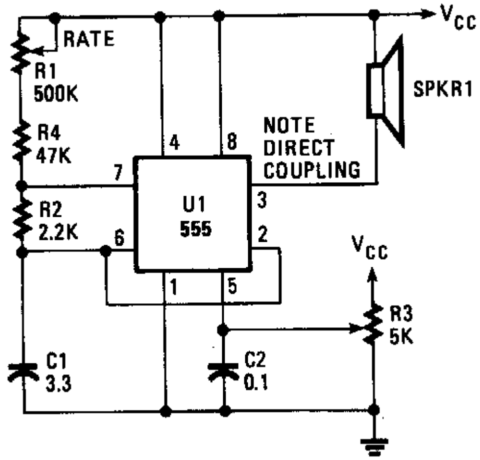
SIMPLE ELECTRONIC METRONOME



POPULAR ELECTRONICS Fig. 50-3

Two complementary transistors form a simple oscillator whose frequency range is from about 0.5 to several Hz. This circuit is useful as a metronome, timer, or pacer for exercise equipment.

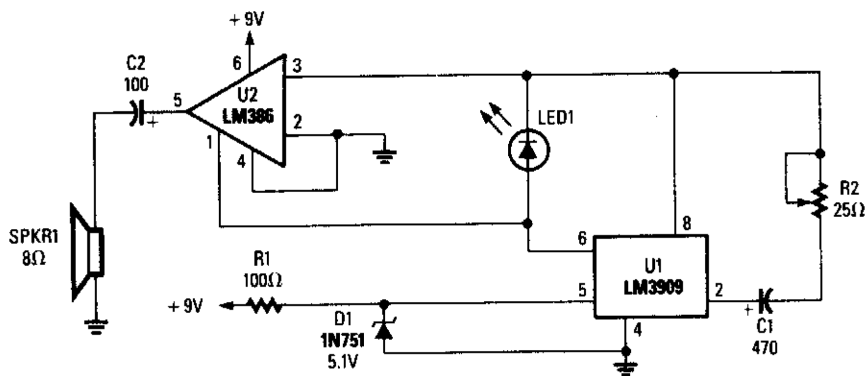
METRONOME II



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS *Fig. 50-4*

This electronomic metronome, using a 555 oscillator/timer, provides 10 to 40 beats per minute. The frequency is controlled by R3.

NOVEL METRONOME



POPULAR ELECTRONICS *Fig. 50-5*

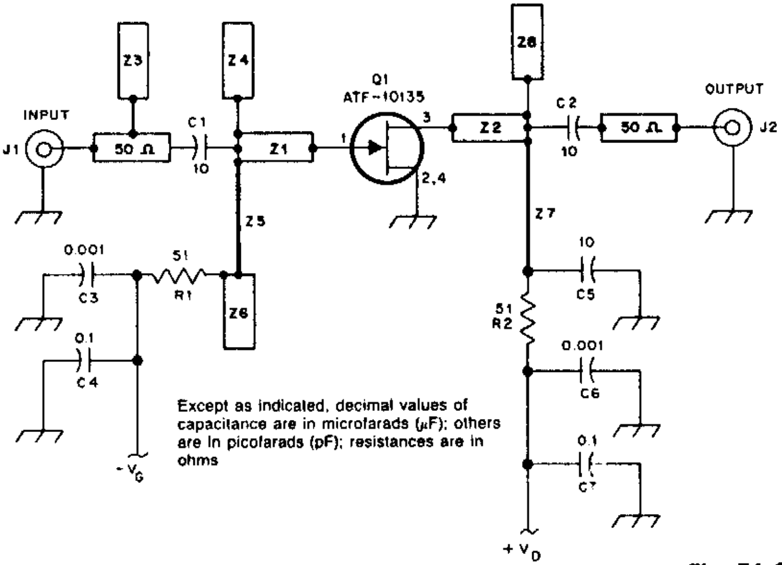
The LM3909 is configured so that the frequency of oscillation is dependent on a single RC timing circuit, which consists of C1 and R2. LED1 discharges capacitor C1 and the resultant pulse is directed into pin 3 as well as pin 1 of the LM386 audio amplifier to externally control that unit, thereby providing adequate volume. The circuit, as it is configured, provides frequency ranges from 57 to 204 beats per minute, and plenty of volume.

51

Microwave Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 671. The figure number in the box of each circuit correlates to the entry in the Sources section.

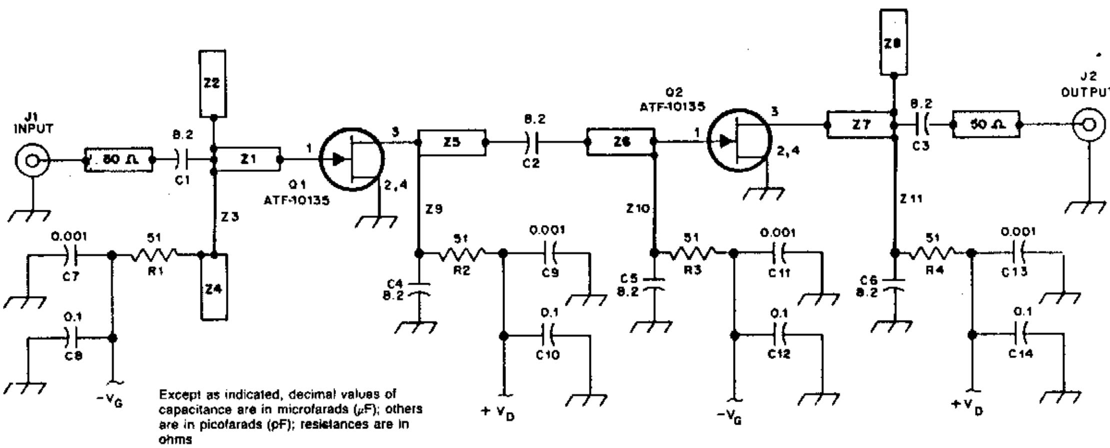
2.3-GHz MICROWAVE PREAMP



QST *Fig. 51-1*

This low-noise amplifier requires no tuning, has a gain of 13 dB, and a typical NF of 0.6 dB at 2.3 GHz.

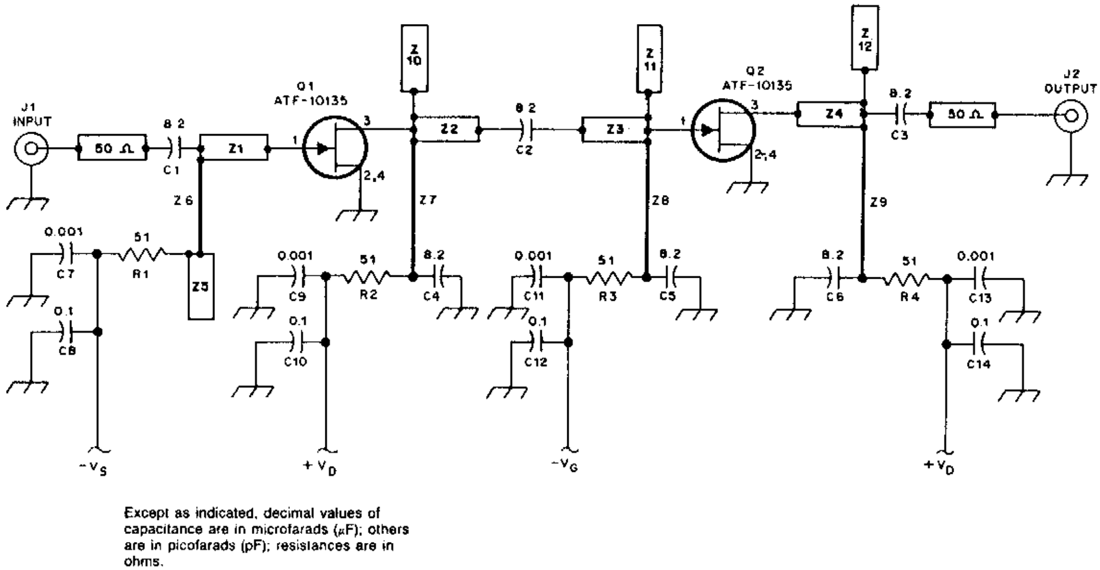
3.4-GHz MICROWAVE PREAMP



QST *Fig. 51-2*

At 3.45 GHz, this 2-stage preamp has a gain of 23 dB (typical) and less than 1 dB NF.

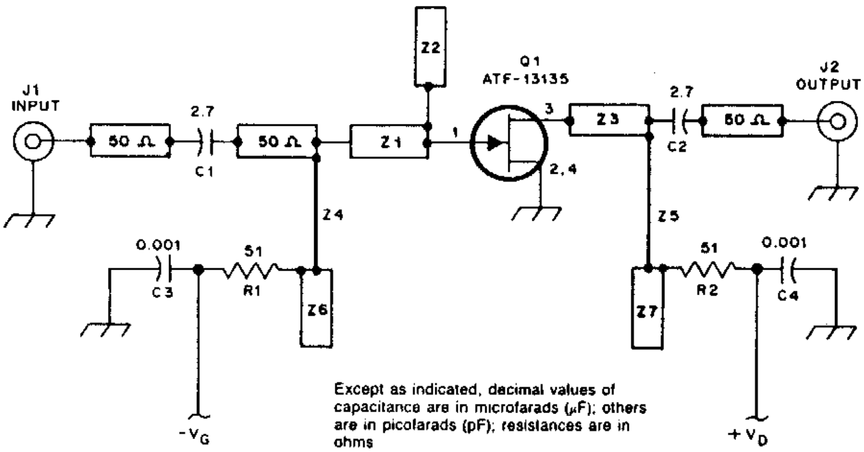
5.7-GHz MICROWAVE AMPLIFIER



QST *Fig. 51-3*

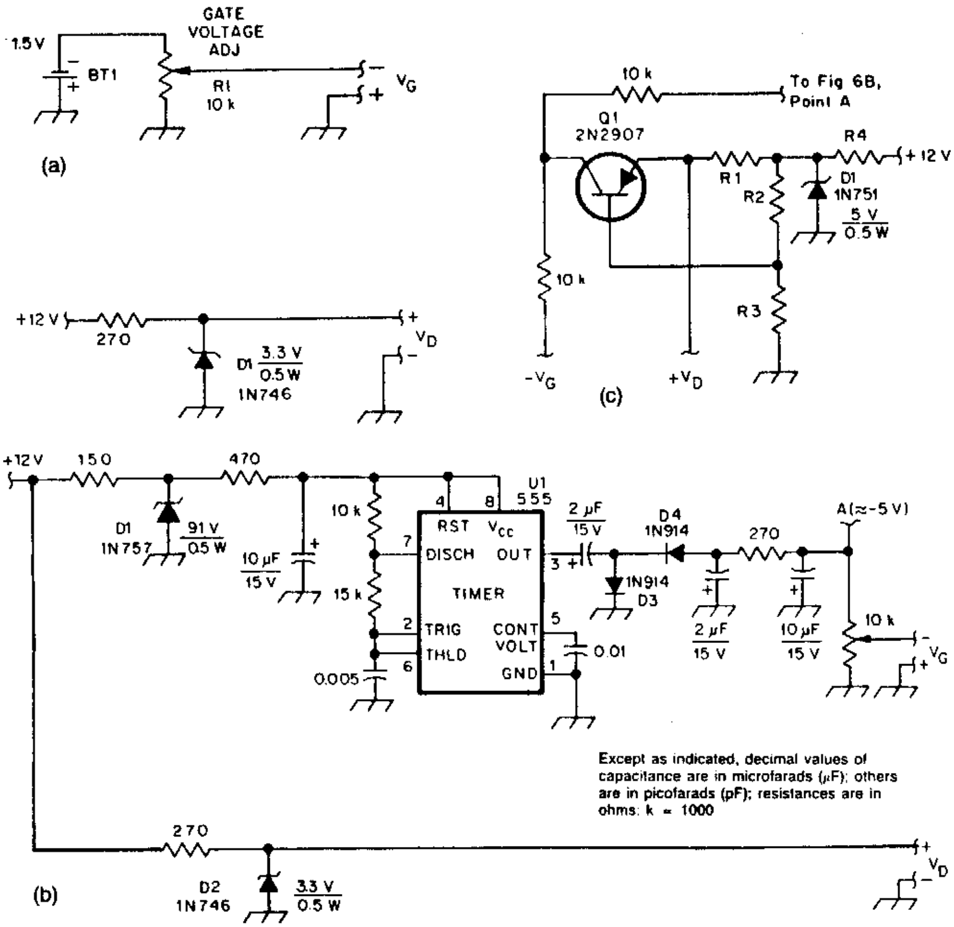
This preamplifier has a typical gain of 17 dB and NF = 1.2 dB or better. If a 0.031″ PC board (with a dielectric constant of 2.2) is used, the reverse side is unetched.

10-GHz SINGLE-STAGE PREAMP



QST *Fig. 51-4*

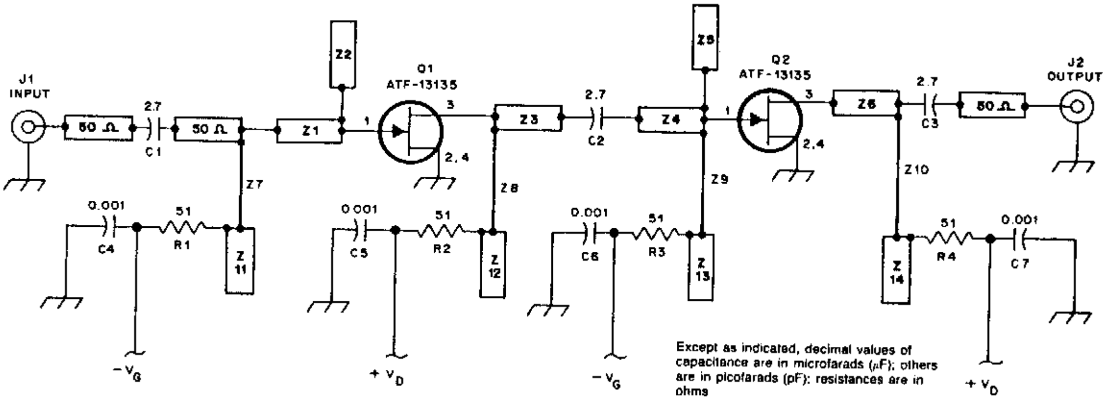
Using a single Avantek ATF 13135 GASFET, this preamplifier has 8-dB gain (typically) and 1.7-dB noise figure. The PC board is 0.031″, doublesided, with $E = 2.2$.



QST | *Fig. 51-5*

These two circuits provide bias for the microwave preamps shown in this text. The circuit in Fig. 51-5(a) is a simple passive supply. Figures 51-5(b) and 51-5(c) are active supplies, with U1 generating a negative supply and Q1 setting the drain voltage and current, independent of GASFET characteristics.

10-GHz 2-STAGE PREAMP



QST

Fig. 51-6

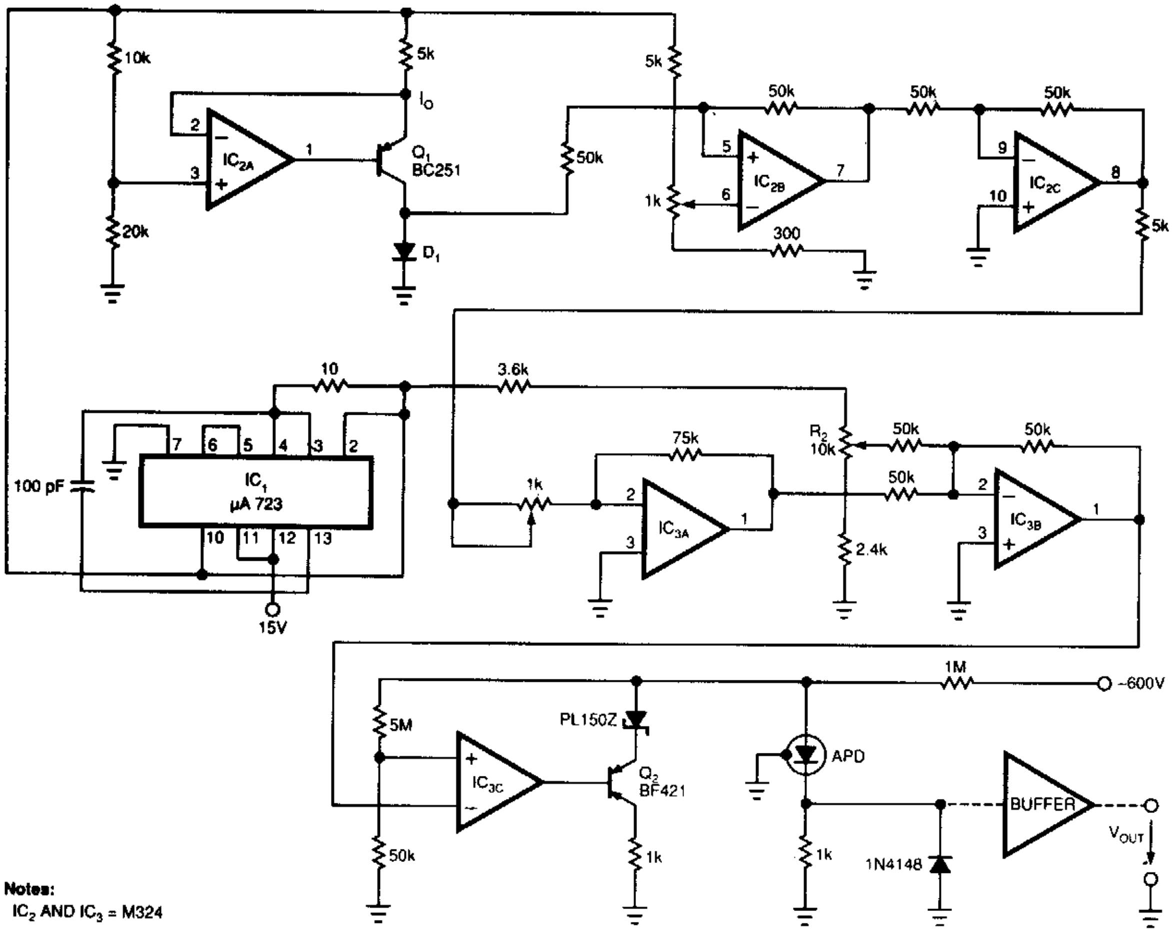
This preamp uses two ATF 13135 stages for typically 17-dB gain and less than 2 dB NF.

52

Miscellaneous Treasures

The sources of the following circuits are contained in the Sources section, which begins on page 671. The figure number in the box of each circuit correlates to the entry in the Sources section.

DIODE SENSOR FOR LASERS



Notes:
IC_2 AND IC_3 = M324

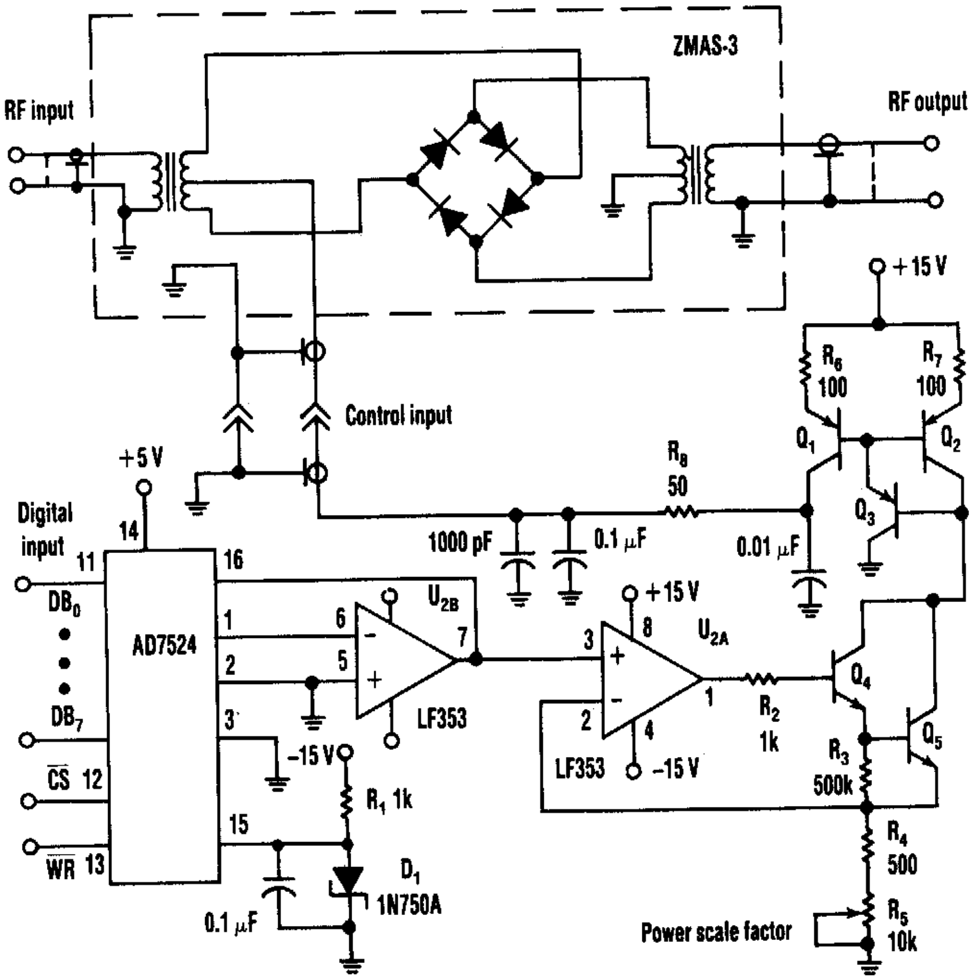
EDN

Fig. 52-1

Laser-receiver circuits must bias their avalanche photo diodes (APD) to achieve optimal gain. Unfortunately, an APD's gain depends on the operating temperature. The circuit controls the operating voltage of an APD over a large temperature range to maintain the gain at the optimal value. The circuit uses D1 as a temperature sensor, thermally matched with the APD.

A voltage regulator, IC1, supplies the necessary reference voltage to the circuit. IC2A and Q1 bias D1 at a constant current. IC2B, IC2C, IC3A/IC3B, and IC3C amplify D1's varying voltage and set Q2 to the optimal-gain corresponding value. Potentiometer R1 controls the amplification over a range of 5 to 15. R2 controls the voltage level, which corresponds to the optimal gain of the APD at 22°C (the temperature is specific to the type of APD). The circuit shown was tested with an RCA C 30954E APD. The tests covered −40 to +70°C and used a semiconductor laser. The laser radiation was transmuted on the APD's active surface in the climatic room via fiberoptic cable. The gain varied by, at most, ±0.2 dB over the entire temperature range.

RF ATTENUATOR

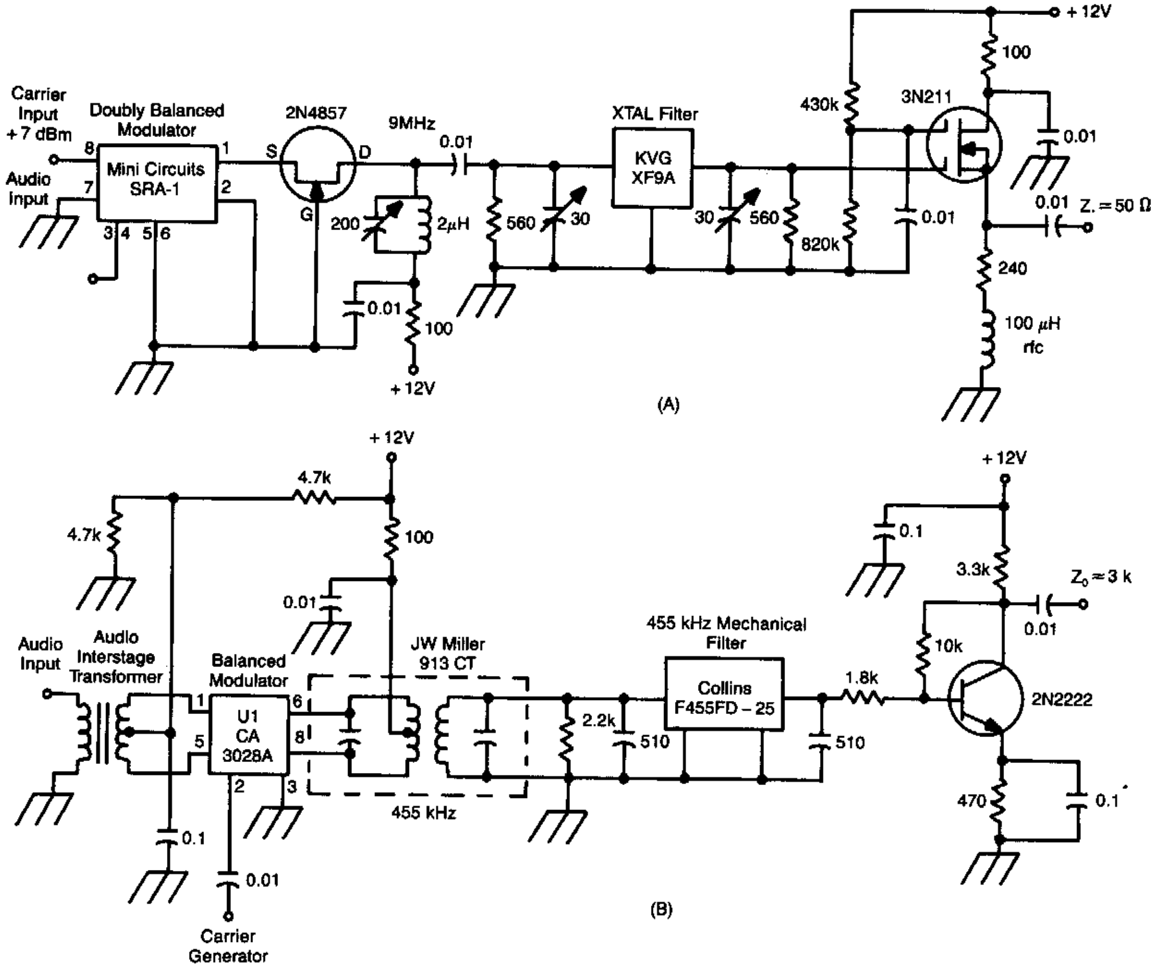


ELECTRONIC DESIGN

Fig. 52-2

A balanced mixer is used as a control element in this circuit. An Analog Devices AD7524 D/A converter drives a voltage-controlled current source using two LF353s and several transistors to control the balanced mixer, a Mini Circuits Lab Z MAS-3.

SSB GENERATORS

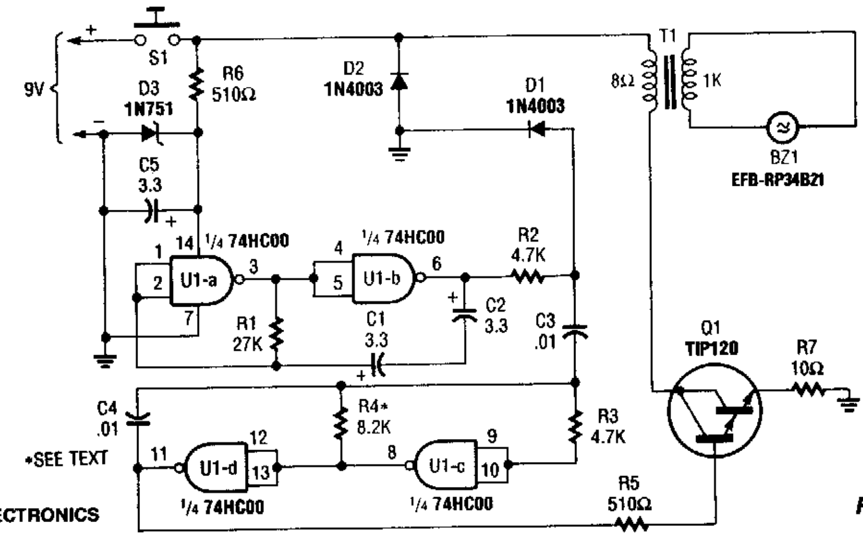


ARRL HANDBOOK

Fig. 52-3

These two circuits are SSB generators. One uses a crystal filter by KVG Electronics at 9 MHz, the other uses a 455-kHz mechanical filter. By feeding the outputs into a mixer, the frequency of the SSB generator can be converted to other frequencies. Keep signal levels low enough so that distortion does not occur.

SONIC DEFENDER



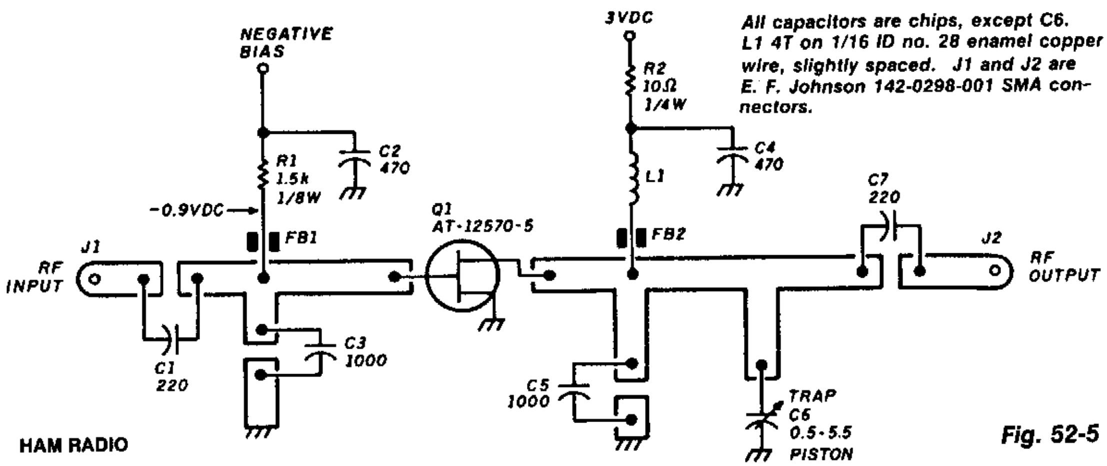
POPULAR ELECTRONICS

Fig. 52-4

This oscillator-driver produces a deafening sound of about 3 kHz, modulated with a 10-Hz warble. BZ1 is a Matsushita EFB-RP34BZ. U1A and B generate the 10-Hz waveform that modulates the 3-kHz tone that is generated by U1C and D. Q1 drives the transducer through a 8-Ω:1-kΩ transformer. R4 might require slight changes (±1 kΩ) to optimize sound intensity.

Warning: Could cause hearing loss if improperly used.

GASFET FREQUENCY DOUBLER

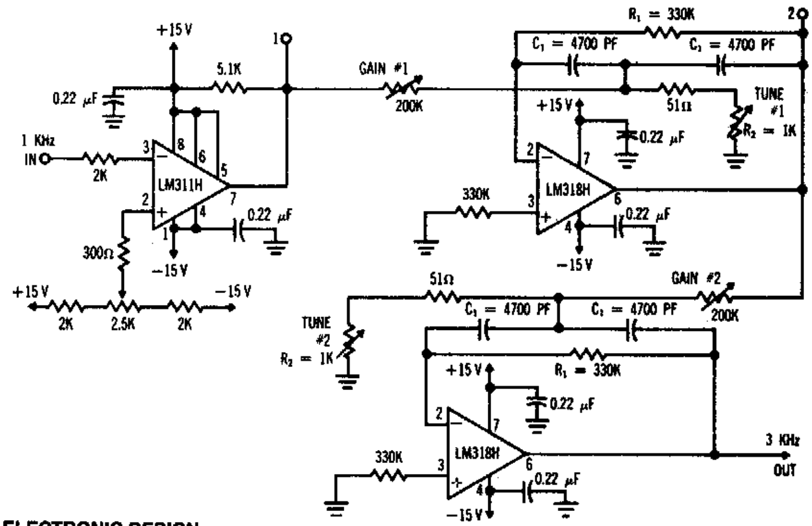


HAM RADIO

Fig. 52-5

This circuit will produce over +10 dBm in the 1 800-3 000-MHz range. Drive power is 7 dBm in the 900-to-1 500-MHz range. The PC board is G-10 Epoxy doublesided. Artwork is shown above, as well as parts placement connectors suitable for these frequencies (such as SMA) should be used. A negative bias supply of 0 to 3 V is required.

LOW-FREQUENCY MULTIPLIER

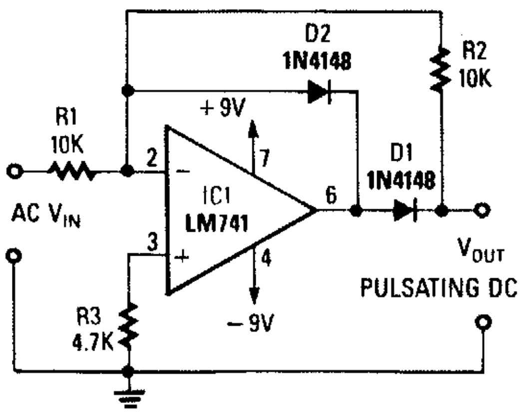


ELECTRONIC DESIGN

Fig. 52-6

This circuit uses a comparator as a Schmitt trigger (311H) and two active bandpass filters (LM318H). 3-kHz output is obtained. Higher harmonics (preferably odd) can be obtained by tuning the active filters to the desired frequency. *N* can be 1, 3, 5, 7, 9, etc. Even harmonics can be produced by substituting a full-wave rectifier or absolute-value circuit for the Schmitt-trigger comparator.

PRECISION HALF-WAVE RECTIFIER

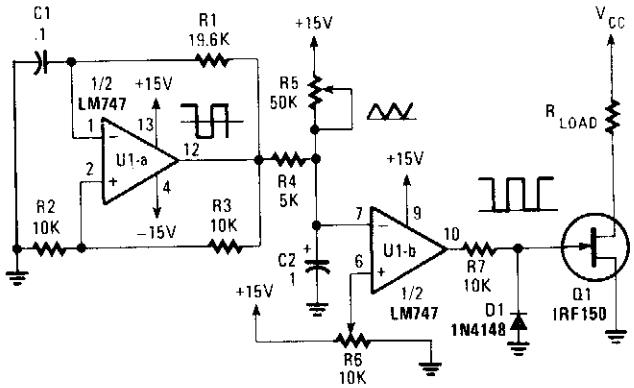


RADIO-ELECTRONICS

Fig. 52-7

An ac input voltage will cause an output that is a half-wave rectified version of the input voltage.

PULSE-WIDTH MODULATOR



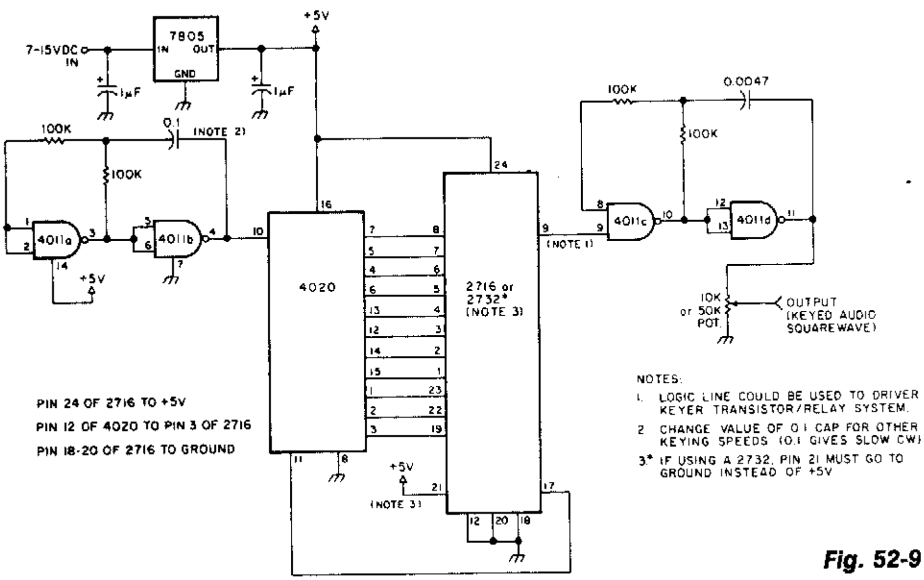
POPULAR ELECTRONICS

Fig. 52-8

This circuit allows the effective power in a load to be controlled by varying the duty cycle of the on/off ratio of load current. No power is dissipated in the switching circuit.

U1A generates a bipolar square wave that is integrated into a triangle by R4 and C2. Reference voltage from R6 is fed to a comparator. The triangle wave on C2 goes to the comparator as well. By varying the reference voltage (R6), the output waveform is a variable width pulse, that drives Q1. R6 controls on/off ratio and therefore load power. R5 sets the offset of the triangle wave across C2.

PROGRAMMABLE IDENTIFIER



FOLLOWING JUMPERS REQUIRED:

PIN 4 OF 4020 TO PIN 6 OF 2716
PIN 5 OF 4020 TO PIN 7 OF 2716
PIN 7 OF 4020 TO GROUND
PIN 10 OF 4020 TO PIN 4 OF 4011
PIN 11 OF 4020 TO PIN 17 OF 2716
PIN 9 OF 2716 TO PIN 9 OF 4011

PIN 24 OF 2716 TO +5V
PIN 12 OF 4020 TO PIN 3 OF 2716
PIN 18-20 OF 2716 TO GROUND

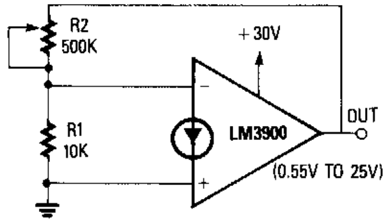
NOTES:
1. LOGIC LINE COULD BE USED TO DRIVER KEYER TRANSISTOR/RELAY SYSTEM.
2. CHANGE VALUE OF 0.1 CAP FOR OTHER KEYING SPEEDS (0.1 GIVES SLOW CW)
3.* IF USING A 2732, PIN 21 MUST GO TO GROUND INSTEAD OF +5V

73 AMATEUR RADIO

Fig. 52-9

Used on an amateur or experimental radio beacon, the above ID circuit will generate any callsign or message programmed into the EPROM. A CD4020 CMOS counter is driven by an RC clock circuit. This addresses the EPROM, and serial data is available at pin 9.

VARIABLE-VOLTAGE REFERENCE SOURCE

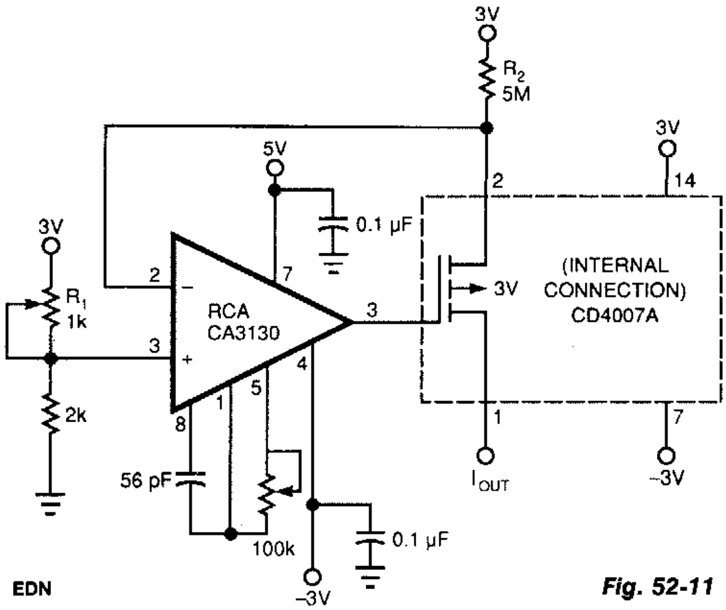


RADIO-ELECTRONICS

Fig. 52-10

The noninverting terminal of the op amp is grounded, and the circuit uses the voltage at the inverting terminal as a reference. Its voltage gain is determined by the R_2/R_1 ratio. When R2 is set at zero, the circuit has unity gain and a 0.55-V output. When R2 is set to the maximum value, the circuit has a gain of 50 and an output of about 25 V. The circuit provides good regulation and can supply output currents of several milliamps. The output voltage however, is not temperature compensated.

0-TO-200-nA CURRENT SOURCE

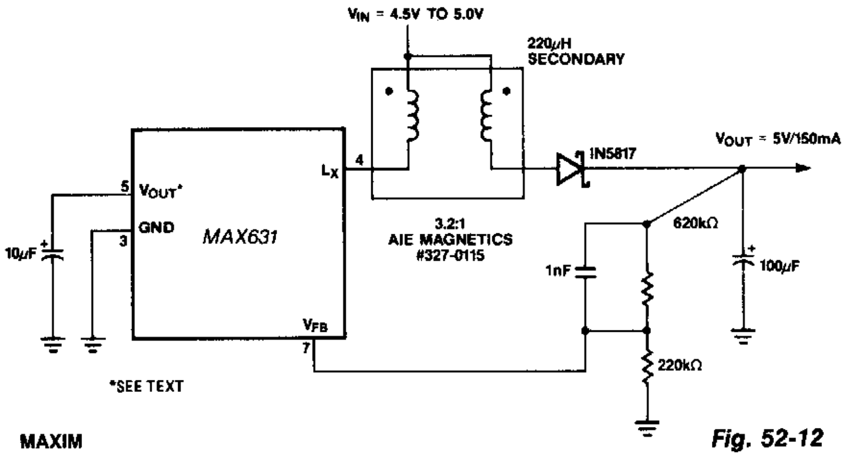


EDN

Fig. 52-11

This circuit uses readily available parts to implement a 0-to-200-nA current source. The circuit borrows a PMOS transistor from the input stage of a DC4007A, which is easier to obtain than a discrete PMOS transistor. The CA3130 op amp operates as a follower so that its positive input sets the current that flows through R2. The MOSFET input stage of this op amp exhibits low-input current. The op amp must be able to produce an output voltage high enough to turn the CD4007A's internal FET off. Thus, the op amp requires a positive supply voltage of 5 V. The circuit presents an output voltage from 0 to 3 V, and R1 controls the amount of output current.

LONG-LINE *IR* DROP-VOLTAGE RECOVERY

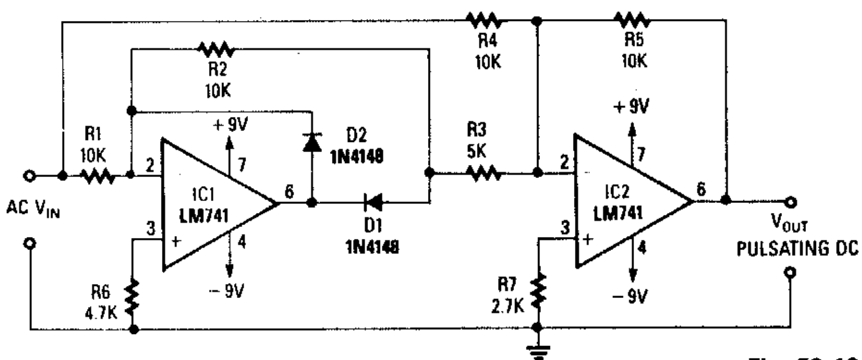


MAXIM

Fig. 52-12

This circuit provides a unique solution to a common system-level power distribution problem: When the supply voltage to a remote board must traverse a long cable, the voltage at the end of the line sometimes drops to unacceptable levels. This +5-V/+5-V converter addresses this by taking the reduced voltage at the end of the supply line and boosting it back to +5 V. This can be especially useful in remote display devices, such as some point-of-sale (POS) terminals, where several meters of cable could separate the terminal from the readout.

PRECISION FULL-WAVE RECTIFIER

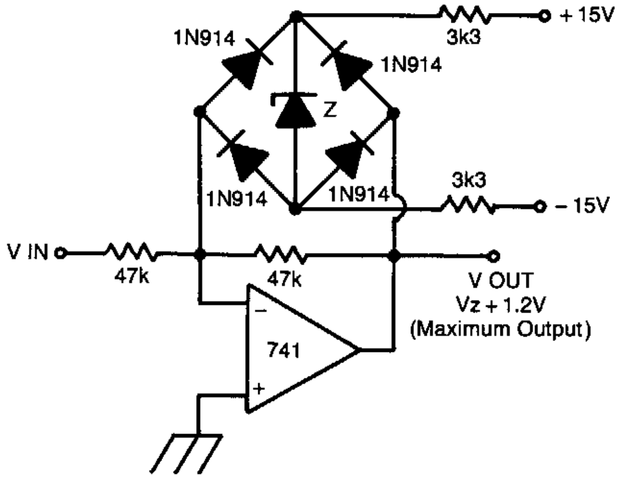


RADIO-ELECTRONICS

Fig. 52-13

Using two op amps, this circuit produces a full-wave rectified version of the input signal. Op amp IC1 inverts the negative-going signal, but because of D2, it stays near zero. IC2 produces a positive-going signal. For positive-going signals, IC1 produces a negative output through D1 to IC2, where it is combined with positive V_{IN} from R4/R5. At the summing junction of IC2, the negative output of IC1 is doubled and inverted via IC2, R3, and R5 to produce $+V_{OUT}$. This is summed with negative output of IC1 to produce $+V_{OUT}$.

FAST SYMMETRICAL ZENER CLIPPER

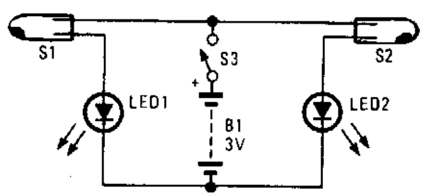


ELECTRONICS TODAY INTERNATIONAL *Fig. 52-14*

The problem with using two zeners back to back in series to get symmetrical clamping is that the knee of the zener characteristics is rather sloppy. Also, charge storage in the zeners causes speed problems and the zeners will have slightly different knee voltages, so the symmetry will not be all that good. This circuit overcomes these problems.

By putting the zener inside a diode bridge, the same zener voltage is always experienced. The voltage errors caused by the diodes are much smaller than those caused by the zener. Also, the charge storage of the bridge is much less. By biasing the zener ON all the time, the knee appears to be much sharper.

ELECTRONIC LEVEL



POPULAR ELECTRONICS *Fig. 52-15*

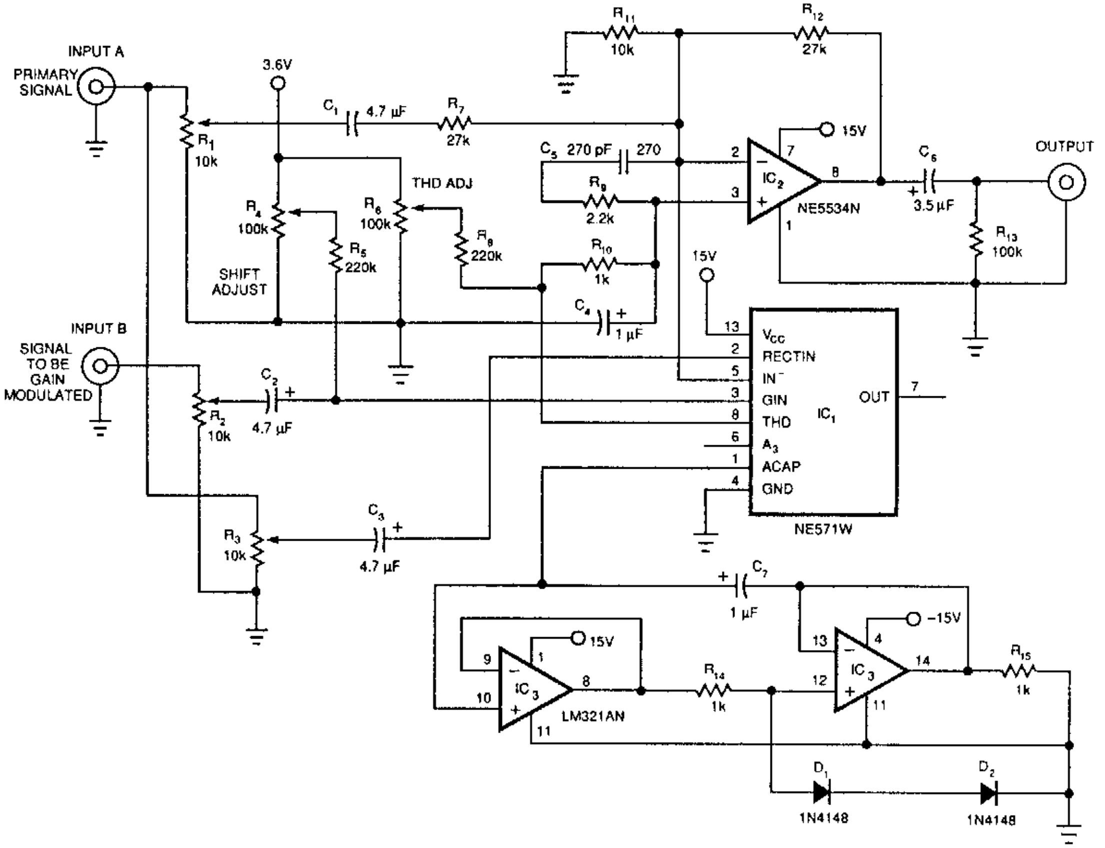
An electronic level can be constructed using two mercury switches mounted on an absolutely flat board, along with the LEDs. If one LED lights, the surface is not level. If both light, the surface is level.

53

Mixers

The sources of the following circuits are contained in the Sources section, which begins on page 671. The figure number in the box of each circuit correlates to the entry in the Sources section.

DYNAMIC AUDIO MIXER



EDN

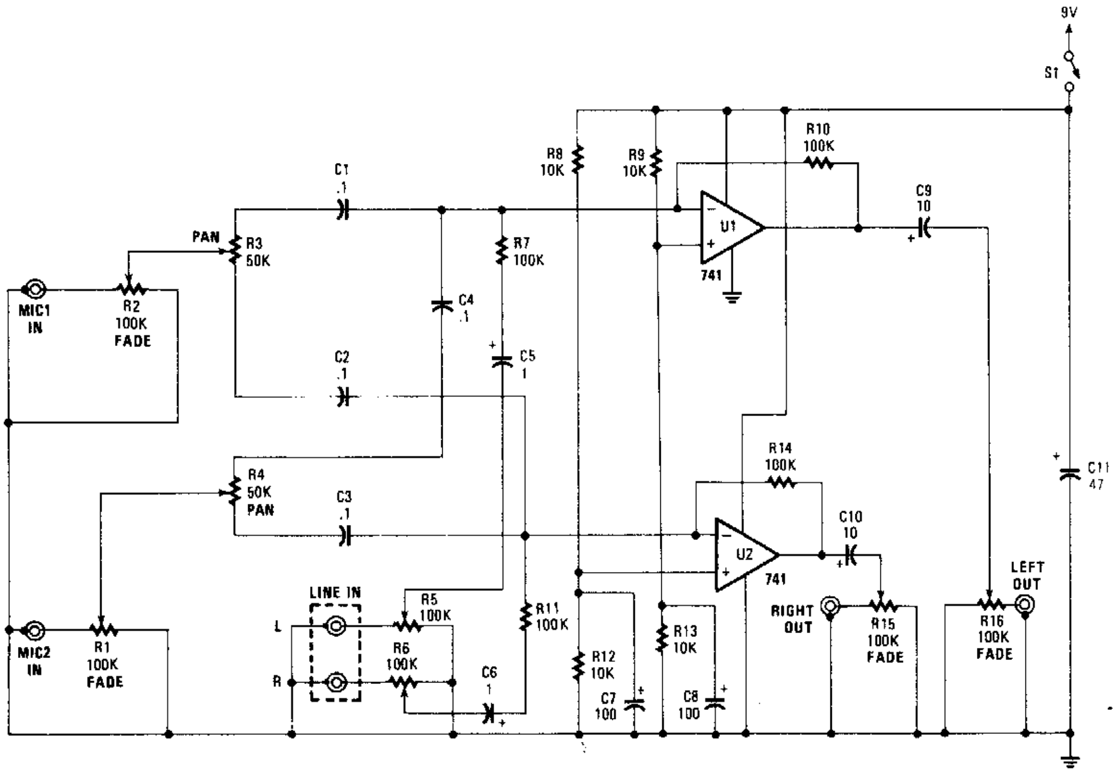
Fig. 53-1

The dynamic mixer combines two audio inputs by adding the primary signal, Input A, to a gain-controlled signal, Input B. The unusual aspect of this circuit is that the average voltage level of Input A controls the gain of Input B.

IC1 has the averaging function and many of the specialized gain blocks that the circuit requires. R1 sets the level of the primary input, Input A, to be passed to the output. R2 governs Input B's level to the modulator, while R3 sets the level of the modulating signal. IC1 can be either an NE571N or an NE570N. The average ac signal at pin 2 controls the amount of signal that shows up at IC1's output, pin 3.

The primary signal gets to IC2, an NE5534N low-noise op amp, via C1 and R7; the gain-modulated secondary signal arrives via pin 5 of IC1. IC2 sums the two signals. Potentiometers R4 and R6 make dc-offset and distortion adjustments, respectively. IC3, C7, R14, R15, D1, and D2 form a filter for IC1.

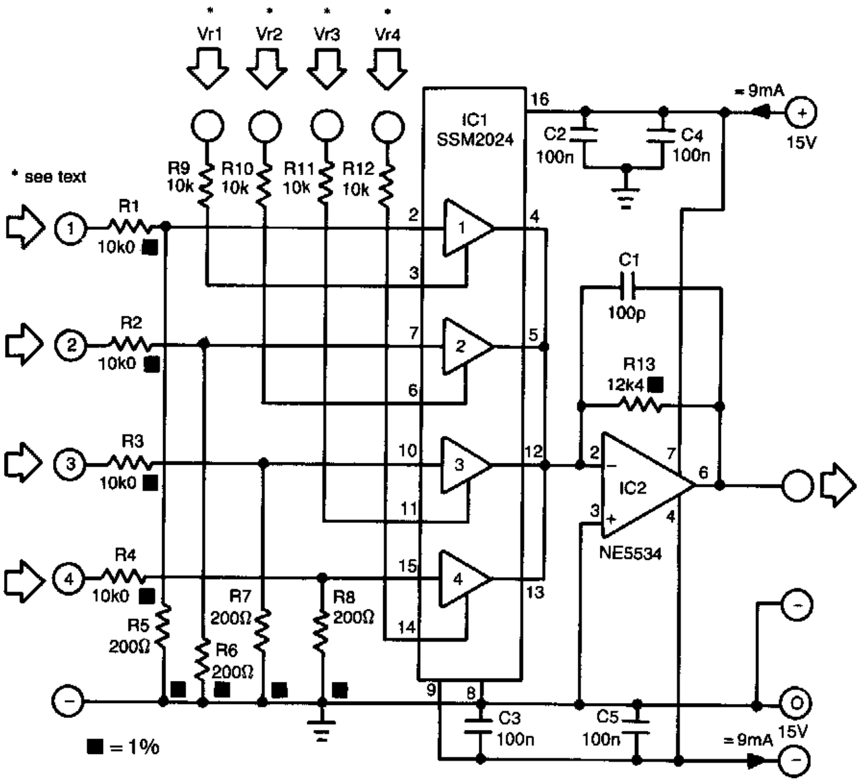
STEREO MIXER WITH PAN CONTROLS



POPULAR ELECTRONICS *Fig. 53-2*

This stereo mixer has two mono mixers and a modification to the microphone inputs. When a microphone is in use, the microphone's output is fed to the microphone input of the circuit. The signal is then run into R1 and R2 (which are used as faders). The signal is then split into two different paths by resistors R3 and R4, with which it is possible to change the place of the microphone inputs within the stereo panorama. The stereo line inputs are for that purpose. Joining the microphone inputs with the output of some other source (such as a tape deck, turntable, etc.), all the signals are fed to the inverting input of an op amp. The output reaches the master-fade potentiometers, which control output level.

4-CHANNEL MIXER



ELEKTOR ELECTRONICS

Fig. 53-3

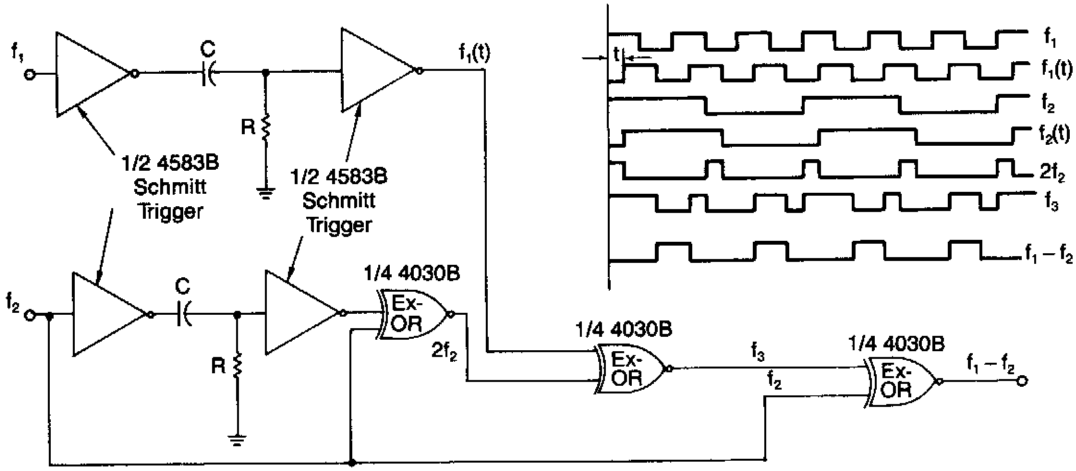
The proposed mixer is designed around four current-driven transconductance amplifiers contained in an SSM2024 from Precision Monolithics. To obtain a low offset and high control rejection, the four inputs should have an impedance to earth of about 200 Ω. These impedances are obtained from resistors R5 through R8, which also form part of a potential divider at each input.

With the values in the diagram, the nominal input signal is 1 V (0 dBV). Distortion at that level is about 1%; at lower levels, it is not more than 0.3%.

The amplification of the current-driven amplifiers (CDAs) is determined by the current fed into the control inputs. These inputs form a virtual earth so that calculating the values of the bias resistors (to transform the inputs into voltage-driven inputs) is fairly simple.

The output currents of the amplifier are summed by simply linking the output pins. The current-to-voltage converter, IC2, translates the combined output currents into an output voltage. The value of R13 ensures that the amplification of IC2 is unity.

DIGITAL MIXER



ELECTRONIC DESIGN

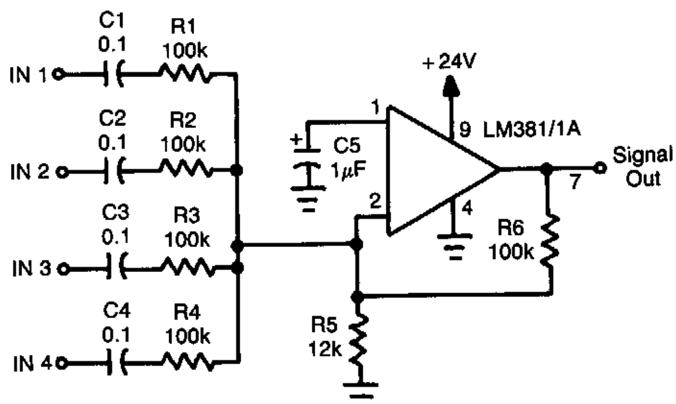
Fig. 53-4

A simple digital mixer, based on two dual-Schmitt triggers (4583B) and three exclusive-OR gates, uses an RC time-delay circuit to permit easy adjustment of the output-signal pulse width. The exclusive-OR gates can also be used separately as a symmetrical frequency doubler.

As shown, a signal passing through the Schmitt triggers is delayed by *t*, a value equal to RC In (V_{tp}/V_{tn}), where V_{tp} and V_{tn} are the positive and negative threshold voltages of the triggers.

To function properly, the same time delay must be introduced to signals f1 and f2. Also, the time delay must be less than 50% of the period of fl. Provided that fl is more than twice the value of f2, the output of the circuit will equal the difference of the two signals (i.e., f1 – f2).

4-INPUT UNITY-GAIN MIXER

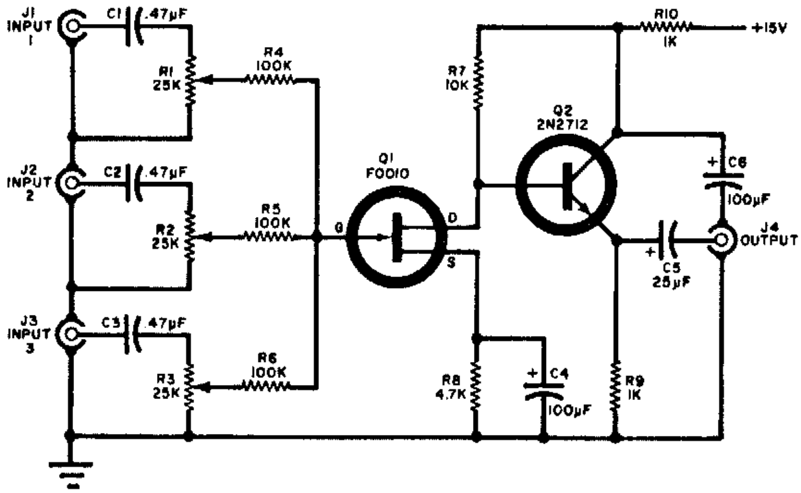


RADIO-ELECTRONICS

Fig. 53-5

An LM381/1A is used as a four-input unity-gain audio mixer. Gain can be increased by decreasing R1 through R4 or increasing R6.

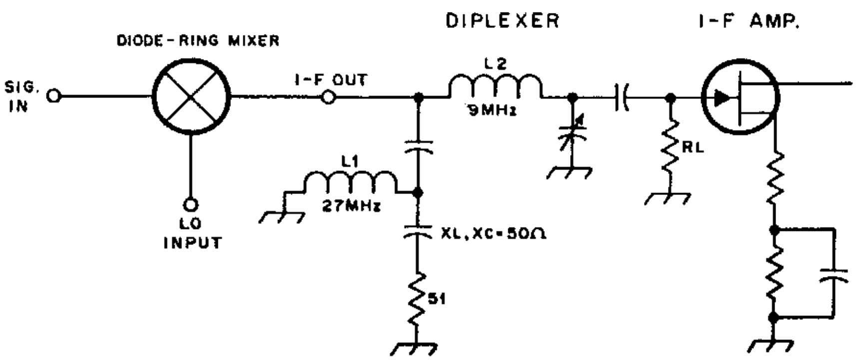
AUDIO MIXER



POPULAR ELECTRONICS *Fig. 53-6*

Three audio circuits are combined in the circuit shown. Each input is coupled to its own level potentiometer (R1, R2, or R3) and they are combined at the gate of FET Q1. The output of Q1 is coupled to the external audio amplifier through emitter-follower Q2 and capacitor C6.

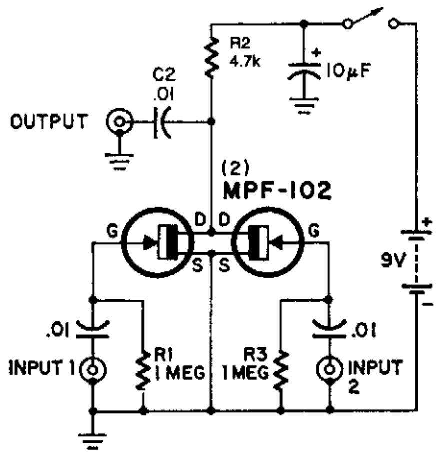
MIXER DIPLEXER



ARRL HANDBOOK *Fig. 53-7*

By inserting a high-pass filter section in the IF lead, this mixer is terminated at all frequencies, besides the IF, for other mixer products, which results in improved IMD. In this example, high-pass filter section L1 and capacitors cut off below 28 MHz. Above this frequency, the mixer is terminated.

SIMPLE UTILITY MIXER



RADIO-ELECTRONICS *Fig. 53-8*

Here's an interesting mixer circuit. With it you can effectively combine signals from audio to high-frequency RF. Also, as a special bonus, this circuit will provide some gain at a low noise figure. The inputs can be of almost any level or impedance, and the output (low-Z) will drive most tuned circuits or transistors.

Basically, the device consists of two similar FET amplifier stages with a common load resistor (R2). Each FET develops a signal across this resistor, a form of cancellation occurs, and a difference signal results.

If you want less gain, try reducing R_2 to 2 200-Ω. This modification will not affect the mixing ability.

54

Model and Hobby Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 671. The figure number in the box of each circuit correlates to the entry in the Sources section.

MODEL TRAIN AND SLOT-CAR CONTROLLER

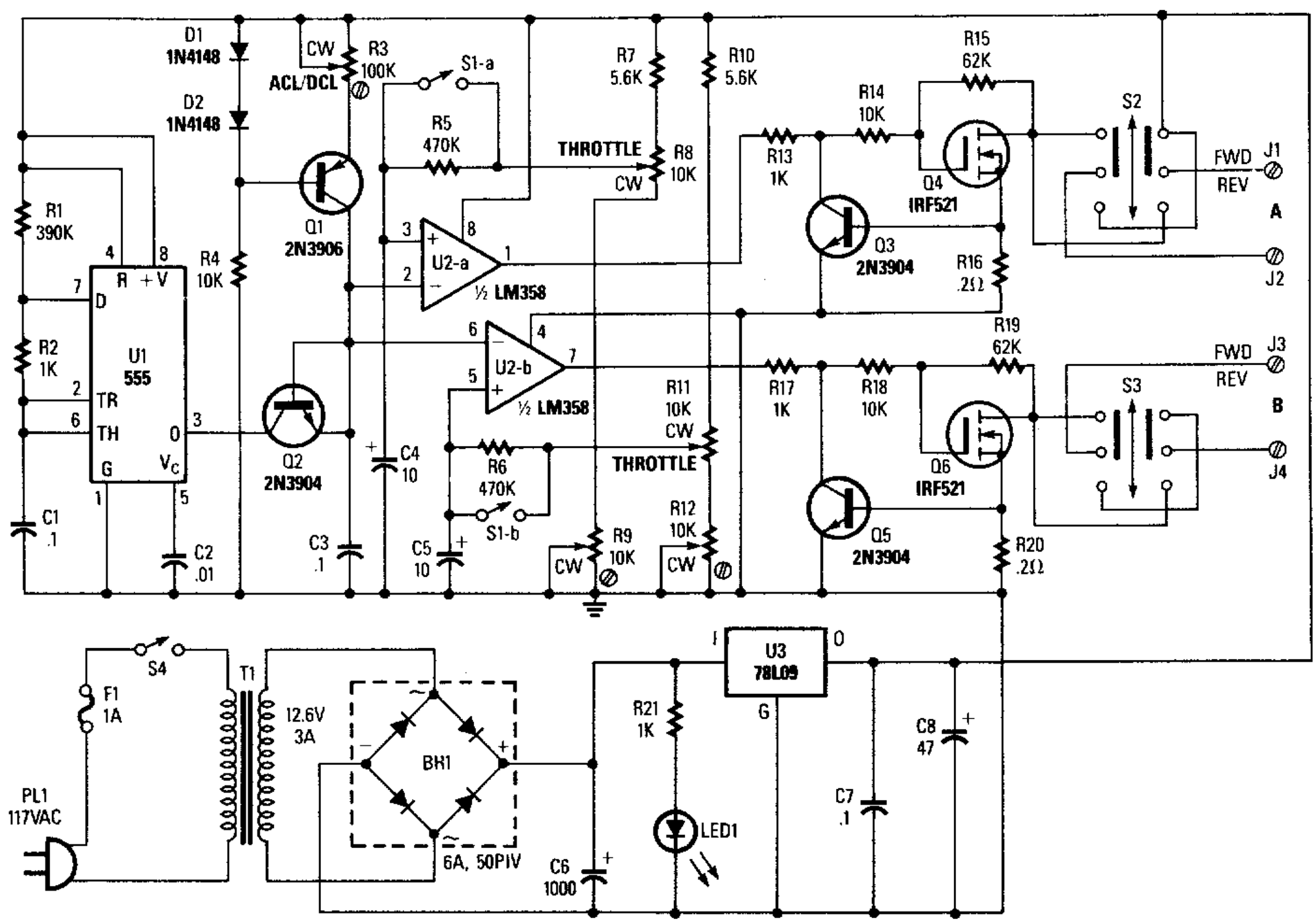


Fig. 54-1

As shown, a 555 timer (U1) is configured as an astable multivibrator (oscillator) with a 400:1 duty cycle and a frequency of 40 Hz.

When power is applied to the circuit, capacitor C1 (connected to pin 6 of U1) is discharged and the output of the 555 (which is used to sink current) is low. Capacitor C1 begins to charge via R1 and R2 toward the positive supply rail. When the charge on C1 reaches about 66% of +V, the output of U1 at pin 3 goes high.

At that point, C1 begins to discharge through R2. When the charge on C1 decreases to about 33% of the supply voltage, the output of U1 returns to the low state, and the cycle is repeated until power is removed from the circuit.

When the output of U1 is low, C3 is discharged into U1 via transistor Q2. When U1 pin 3 goes high, C3 charges through a current source that consists of D1, D2, R3, R4, and Q1. The charge/discharge cycling of C3 produces a stream of pulses that are fed to the inverting inputs of U2A and U2B (an LM358 dual op amp). Two voltage-divider networks (R7, R8, R9, and R10, R11, R12) set the reference voltage that is applied to the noninverting inputs of U1A and U1B at pins 3 and 5.

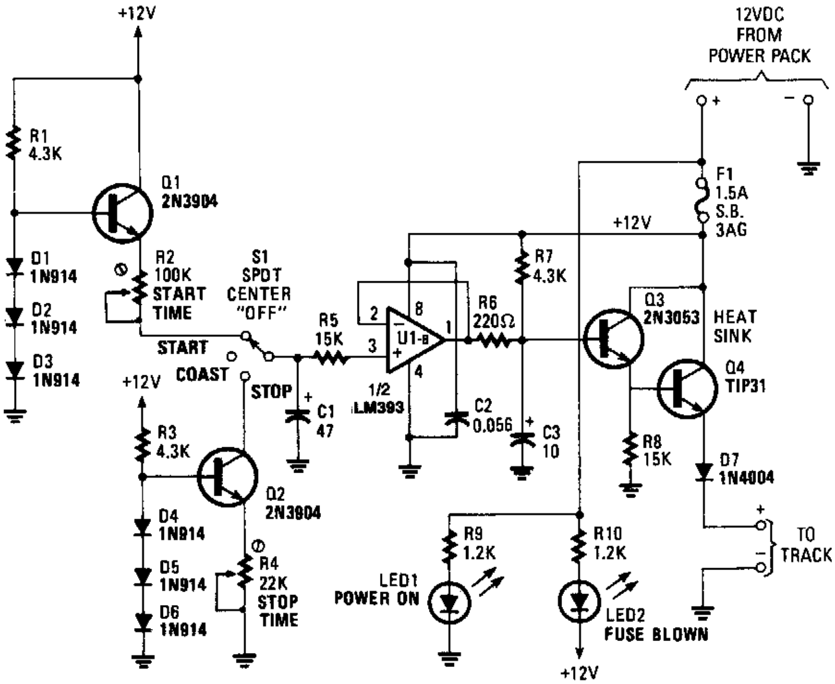
Potentiometers R9 and R12 set the low-level duty cycle (5 to 10%) of U1A and U1B. They are adjusted so that the train headlights glow, but the motor hums only slightly. Potentiometer R3 adjusts the ramp rate of C3 for 100% duty cycle at the full throttle setting. A double-pole, single-throw switch (S1A and S1B) is used to place R3/C4 and R4/C5 in the circuit.

The R5/C4 and R6/C5 combinations cause the reference voltages presented to the noninverting inputs to U2A and U2B to change very slowly when the throttle is turned up and down. When the ACL/DCL switch is turned off, the resistance of the throttle-divider networks are much smaller than those of R5 and R6, so the reference voltages on C4/C5 change "instantly" to the new throttle setting.

The output drivers consist of resistors R13 and R15, and transistors Q3 and Q4 for output "A"; and resistors R17 to R20, and transistors Q5 and Q6 for output "B." Components R13/R16/Q3 and R17/R20/Q5 limit the output drive currents of Q4 and Q6 to about 3 A each. Resistors R14/R15 and R18/R19 turn on Q4 and Q5, respectively, before the breakover voltage is reached to prevent damage to the output drivers and dissipate the energy that is stored in an inductive field (such as in a motor).

The power supply delivers 18 V to the track. Voltage regulator U3 (a 78L09 9-V, 100-mA voltage regulator) supplies power to the control circuits.

MODEL TRAIN THROTTLE CONTROL



POPULAR ELECTRONICS *Fig. 54-2*

What makes this control unique is its momentum feature, which adds a degree of realism. The circuit will operate well for trains that draw up to 1 A at 15 V. None of the components are critical.

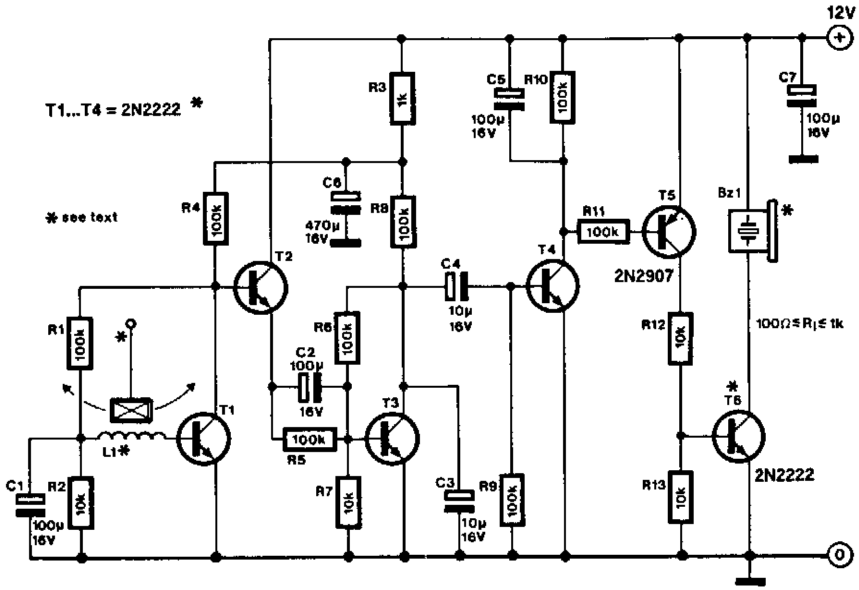
In the start mode, current source Q1 charges capacitor C1. The charge current and start-up time are adjusted by resistor R2. In the stop mode, current-sink Q2 discharges capacitor C1. The discharge current and stop time are set by resistor R4. In the coast mode, op amp U1 draws very little current from C1, so the speed will remain nearly constant for some time, and then gradually decrease. Transistors Q3 and Q4 form a Darlington emitter-follower to amplify the output of U1. Diode D7 reduces the output by about 0.8 V. Another diode could be added in series to decrease the output to 0 in the stop mode.

55

Motion and Proximity Detectors

The sources of the following circuits are contained in the Sources section, which begins on page 671. The figure number in the box of each circuit correlates to the entry in the Sources section.

SENSITIVE LOW-CURRENT-DRAIN MOTION DETECTOR



ELEKTOR ELECTRONICS

Fig. 55-1

This highly sensitive movement detector is designed from bipolar transistors and draws a current of only 0.3 mA during quiescent operation. It is intended primarily as a protection device, but it can also be used in certain games.

The principle is simple: a magnet is suspended by a thin thread 20 to 30 mm long, and a few millimeters above the coil of a relay (whose contacts are not used). Even a minute movement of the protected object will disturb the magnet. The resulting changes in the magnetic field above the relay coil will induce a tiny varying voltage across the coil.

The first stage consists of a common emitter design with automatic regulation. The collector resistors and the resistors in the regulation bridge have unusually high values.

Feedback from the bridge ensures stability of operation for T1. Each increase in collector voltage will be opposed by an increase in base-emitter current. Conversely, each reduction in collector voltage will be opposed by a decrease in base-emitter current. Consequently, the collector voltage will stabilize at a value that corresponds to a base voltage of about 0.6 V. Capacitor C1 delays the immediate effect of the feedback when the collector voltage changes rapidly.

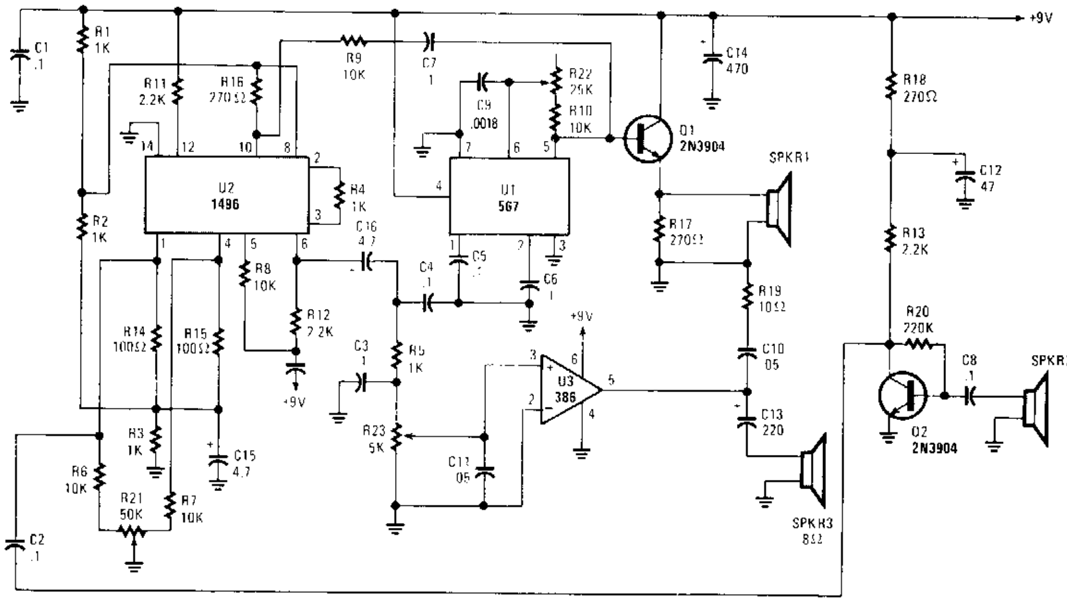
The small varying voltage induced in the relay coil is magnified appreciably by T1 because C1 prevents automatic regulation. The output impedance of the first stage is very high, which is, of course, the price to be paid for low consumption. It would not make sense to follow this stage by one with a low output impedance, because this would adversely affect the overall amplification. Because of that, T1 is followed by an emitter-follower, T2, which provides the coupling between T1 and T3. Resistor R5 allows a partial discharge of C2 if T2 is switched off by a reduction in the output of T1. Because this resistor, as a result of the low-consumption requirement, has a high value, the circuit will attain its maximum sensitivity 10 seconds after the last movement detection. This is the time that is required for the charge on capacitor C2 to stabilize.

The detection proper is carried out by T4, which switches on when the voltage variations in the amplifier, passed on by C4, reach a level of 0.6 V. Saturating T4 leads to the instant charging of C5. This capacitor will discharge partly via R10 and R11 to the base of T5 when T4 switches off again. When C5 discharges, T5 is on, which will make T6 conduct. This in turn will actuate a load, for instance, a buzzer, in the collector circuit of T6.

The sensitivity of the detector depends to a large extent on the distance between the magnet and relay and the length of the "pendulum."

If the circuit is powered by a battery, there is a little problem: batteries have large internal resistances. Thus, a supply voltage can vary by some tenths of a volt if a sudden, large current is drawn. If the buzzer has stopped after a detection, such a situation can retrigger the circuit and cause undesired oscillations. To prevent this happening, the supply of the amplifier stage is decoupled by R3 and C6.

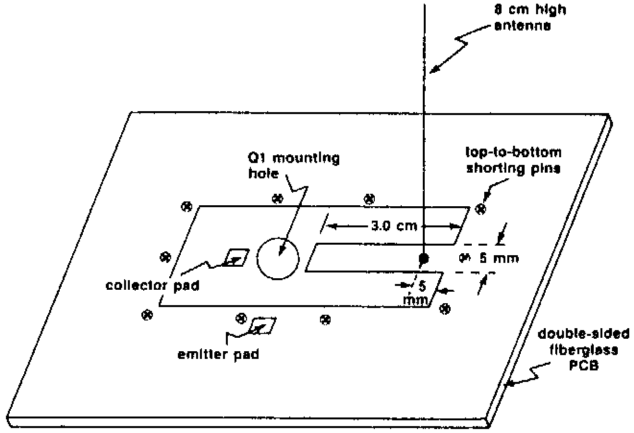
ACOUSTIC DOPPLER MOTION DETECTOR



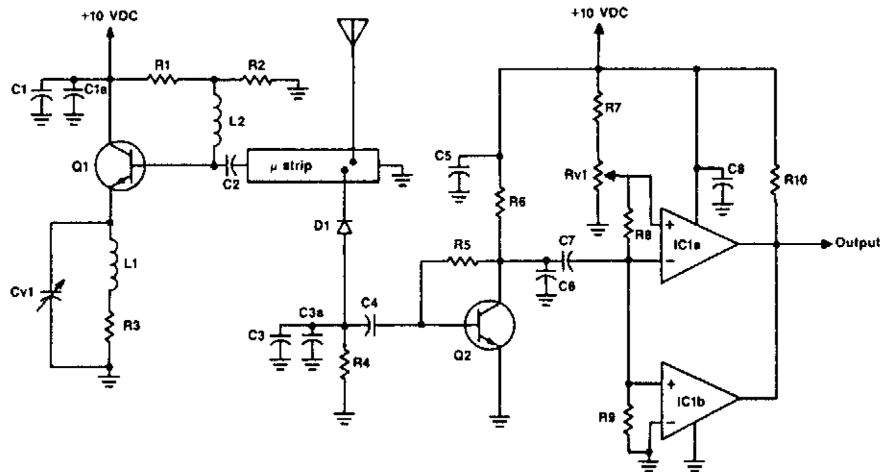
POPULAR ELECTRONICS ***Fig. 55-2***

A high-frequency audio signal (15 to 25 kHz) generated by U1 is fed to buffer Q1 and SPKR1. A portion is fed to balanced mixer U2. Received audio picked up by SPKR2 (used as a microphone) is amplified by Q2 and fed to U2. When sound is reflected from a moving object, the Doppler effect will cause an apparent shift in frequency. U2 produces a signal equal to the frequency difference. This is coupled via C16 and gain control R23 to amplifier U3, where the beat note is heard in SPKR3.

UHF MOTION DETECTOR



Construction layout.



Parts List

				D1	ISS97 or other Schottky type
R1	3.9 k	C1	1 nF	Q1	MRF961
R2, 4	1 k	C1a	470 uF	Q2	BC546
R3	100	C2	47 nF	Cv1	2-7 pF miniature
R5	2.2 M	C3a	1 uF	IC1	LM339 comparator
R6	6.8 k	C4	22 uF	L1	5 turns 0.86 mm wire on
R7	100 k	C5, 8	100 uF		3.5 mm core
R8, 9, 10	22 k	C6	100 nF	L2	4 turns 0.86 mm wire on
Rv1	1 k	C7	10 uF		3.5 mm core

RF DESIGN

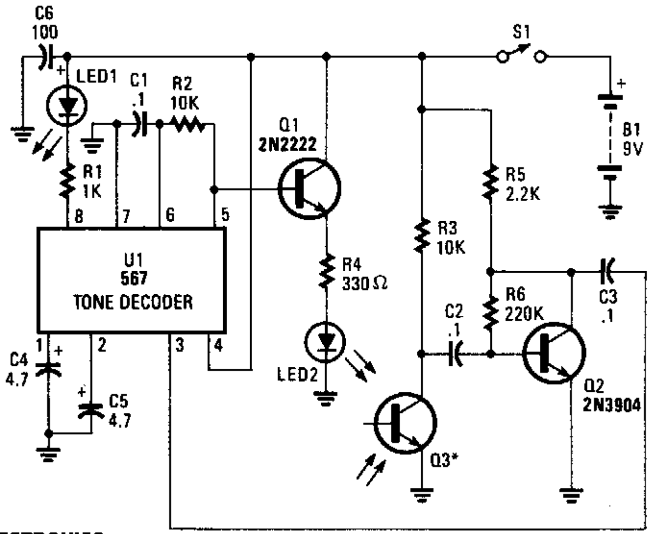
Fig. 55-3

Circuit diagram.

The UHF motion detector operates on the Doppler radar principle. Q1 is an oscillator that creates a radiated signal. An object in the radiated field reflects some of this energy back to the detector. If the object is moving, the reflected signal will have a different frequency because of Doppler shift.

Q1 is an oscillator coupled to a small (8-cm) antenna. This antenna also receives the reflected signal. D1 acts as a mixer and produces a beat note of frequency that is equal to the difference in reflected and radiated signal. Q2 amplifies this signal and couples it to comparator/detector IC1A and IC1B. The output can continue on to an alarm, relay, lamp, etc.

IR REFLECTION PROXIMITY SWITCH

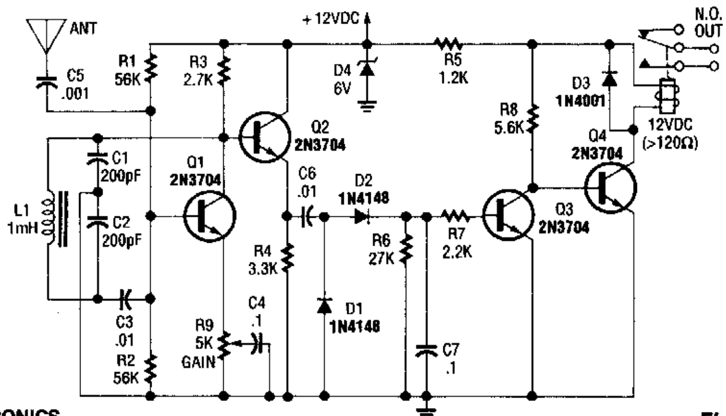


POPULAR ELECTRONICS

Fig. 55-4

IR radiation from LED2 (modulated by a 1-kHz wave) is keyed by U1, and Q1 is radiated. Reflected IR energy is picked up by Q3, and the audio signal from Q3 is amplified by Q2 and sent to the decoder. The LED1 lights to indicate presence of reflected IR. LED1 can be the input of an isolator so that a triac or SCR can be controlled.

RELAY OUTPUT PROXIMITY SENSOR

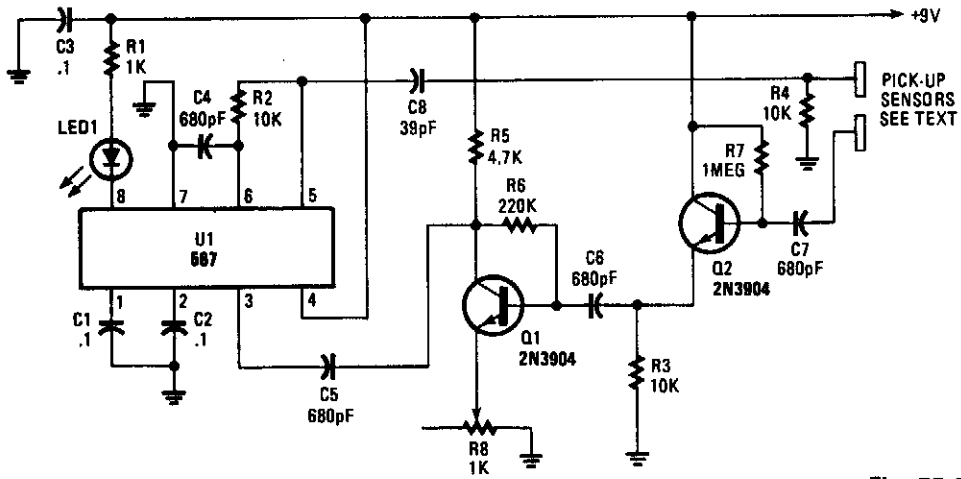


RADIO-ELECTRONICS

Fig. 55-5

Q1 is used as an oscillator around 300 kHz. R9 is set so that the oscillator just begins to run. An object near the antenna will load the circuit down, and stop the oscillations. This is detected by buffer Q2, diodes D1 and D2, and this activates relay driver Q4, which operates the relay.

PROXIMITY DETECTOR



POPULAR ELECTRONICS

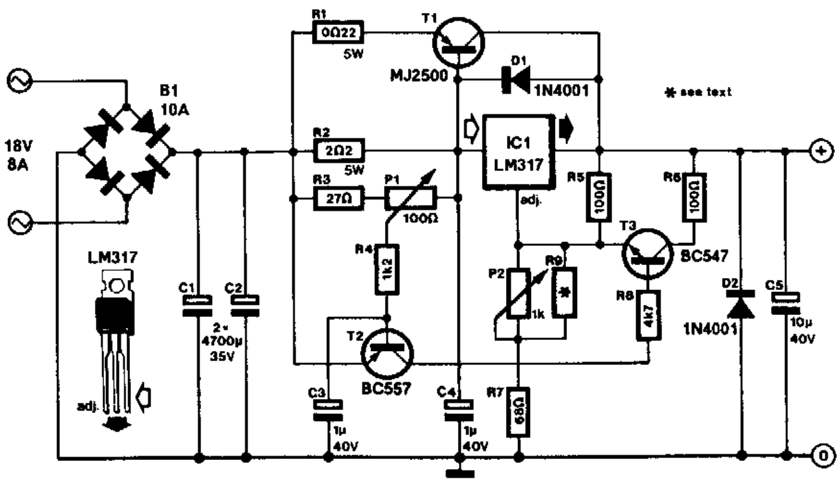
Fig. 55-6

In this proximity detector an NE567 tone decoder provides a signal of about 100 kHz that is fed to one sensor. The sensors are copper or aluminum wires or plates, or any other suitable conductive material. When another object is near the sensors, it causes an increase in capacitance between the sensors. Q1 and Q2 amplify the signal and feed it to U1. Notice that C8 and R4 phase shift the VCO signal from the NE567 so that U1 can detect its own output signal.

56

Motor Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 672. The figure number in the box of each circuit correlates to the entry in the Sources section.



ELEKTOR ELECTRONICS

Fig. 56-1

This circuit is intended as a revolution control for small dc motors as fitted, for instance, in small electric drills (such as used for precision engineering and for drilling boards, among others). The behavior of these motors, which are normally permanent magnet types, is comparable to that of independently powered motors.

In theory, the rpm of these motors depends solely on the applied voltage. The motor adjusts its rpm until the counter emf generated in its coils is equal to the applied voltage. There is, unfortunately, a drop across the internal resistance of the motor, which causes the rpm to drop in relation to the load. In other words, the larger the load, the larger the drop across the internal resistance and the lower the rpm.

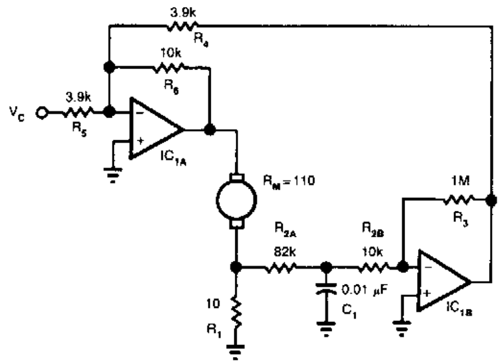
This circuit provides a kind of compensation for the internal resistance of the motor: when the current drawn by the motor rises, the supply voltage is increased automatically to counter the fall in rpm.

The circuit is based on an enhanced voltage regulator that consists of IC1 and T1, which provides a reasonably large output current (even small drills draw 2-to-5 A). The "onset" supply voltage, and thus the rpm, is set by P2. Because of emitter resistance R1, the currents through IC1 and T1 will be related to one another in the ratio that is determined by R1 and R2. Owing to this arrangement, the internal short-circuit protection of IC1 will also, indirectly, provide some protection to T1.

As soon as the current drawn exceeds a certain value, T2 will be switched on. This results in a base current for T3 so that R5 is in parallel (more or less) with R6. This arrangement automatically raises the output voltage to counter a threatened drop in rpm. The moment at which this action occurs is set by P1, so this circuit can be adapted pretty precisely to the motor used.

If only very small motors are likely to be used, the power supply (transformer and bridge rectifier) can be rated more conservatively. As a guide, the current in the transformer secondary should be about 1.5 times the maximum dc output current.

TACHOMETERLESS MOTOR SPEED CONTROL



EDN

Fig. 56-2

This provides bidirectional speed regulation for small motors and requires no tachometer. The voltage that summing amplifier IC1A applies to the motor's windings equals:

$$\left(V_C + R_1\left(\frac{R_3}{R_{2A}+R_{2B}}\right)\right)I_M\left(\frac{R_6}{R_5}\right)$$

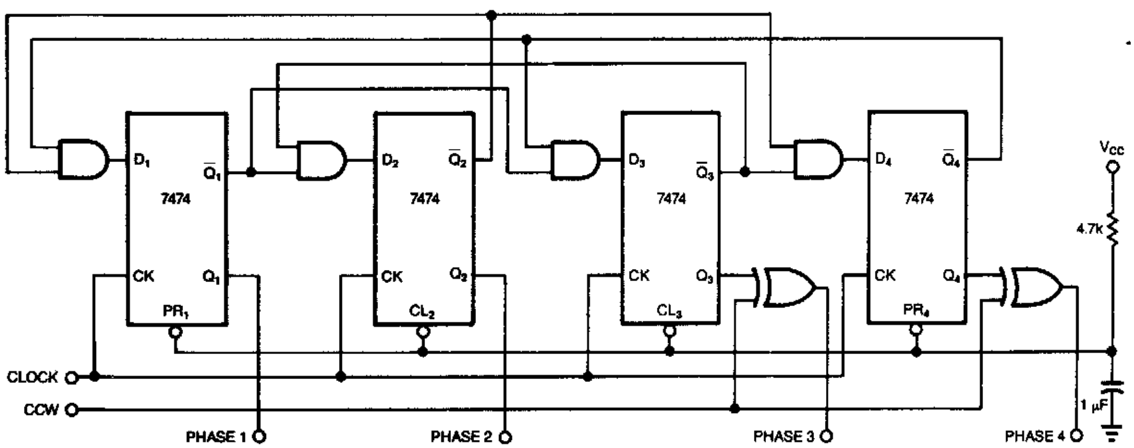
where V_C is the command voltage and I_M is the motor current.

If you set the motor's winding resistance and brush resistance (R_M) equal to:

$$R_1\left(\frac{R_3}{R_{2A}+R_{2B}}\right)$$

the command voltage will be proportional to the motor winding's counter emf. C1 provides compensation. Set R1's value so that it equals 5 to 10% of RM's value. You can generally find RM's value in a motor's spec sheet.

HALF-STEP DRIVE STEPPER MOTOR

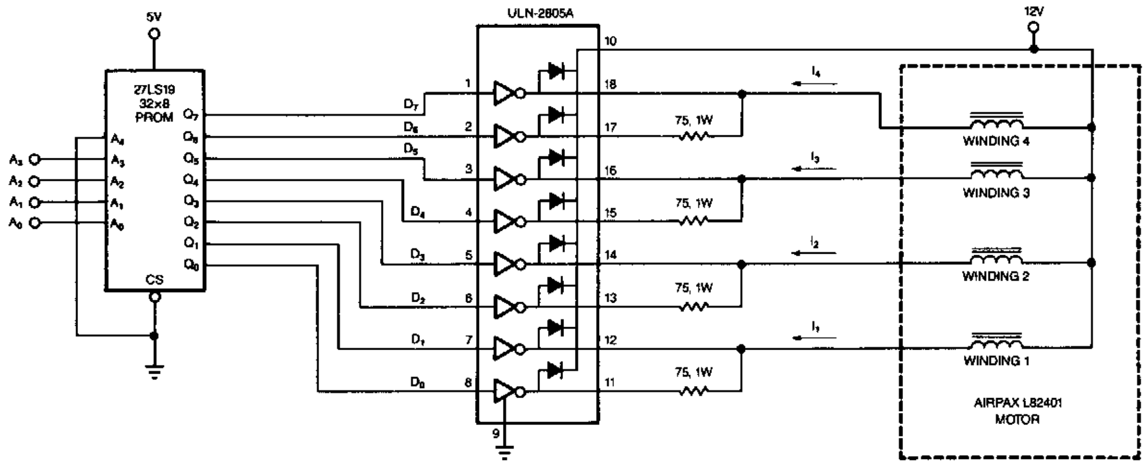


EDN

Fig. 56-3

Clock pulses are converted to the switching sequence necessary for a four-phase stepper through 400 half steps of 0.9° each.

QUARTER-STEP STEPPER MOTOR DRIVER

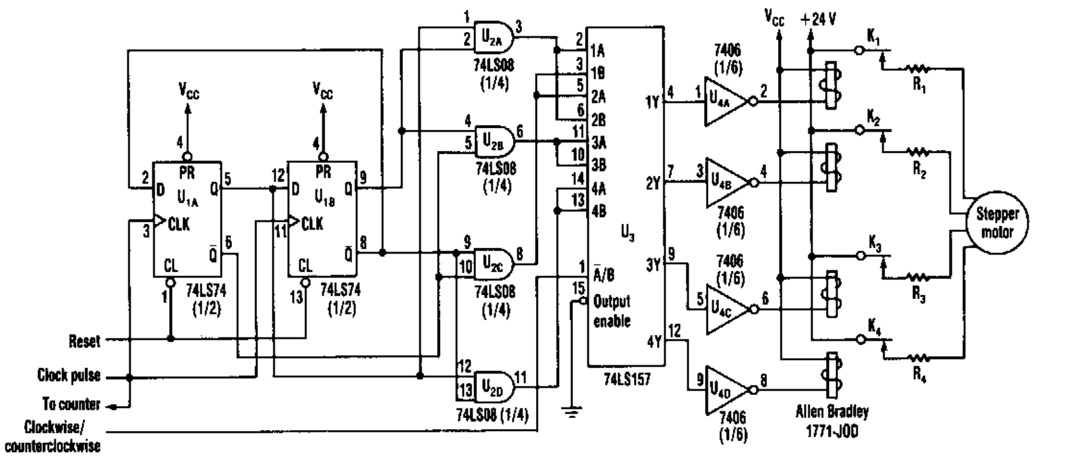


EDN

Fig. 56-4

Using a counter and a PROM, this circuit drives the stepper windings with two levels of current.

STEPPER MOTOR SPEED AND DIRECTION CONTROLLER

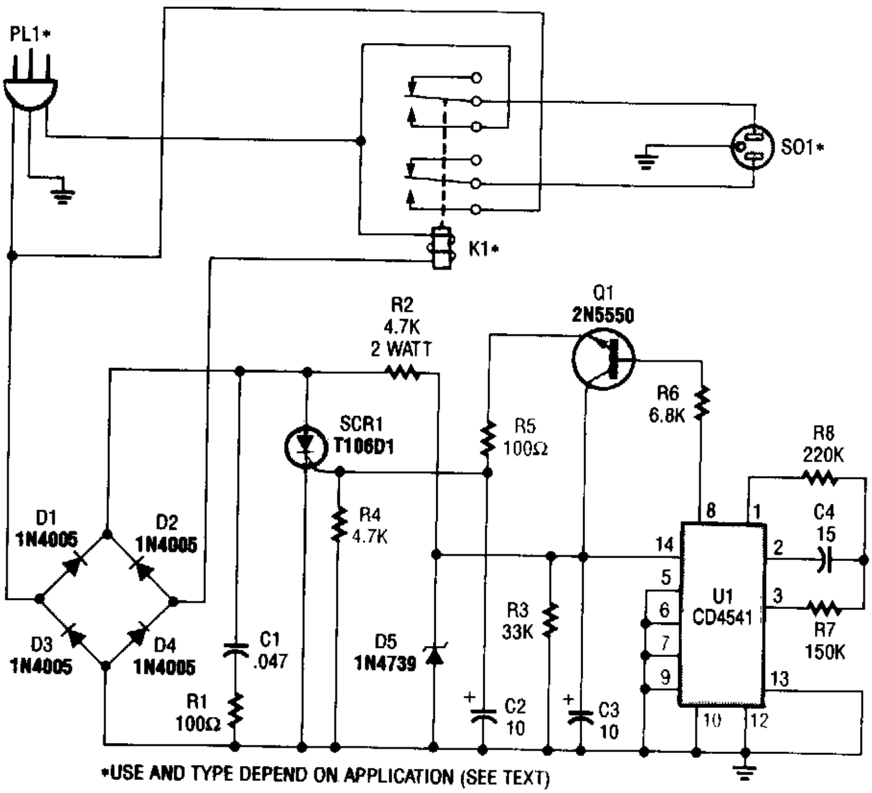


EDN

Fig. 56-5

This new circuit uses four chips, with an option of using just three (the flip-flops and AND gates can be combined). The rate of clock pulses determines the motor's rpm. Switching transistors can replace the relays to increase the circuit's efficiency. This circuit drives a stepper motor whose speed depends on clock rate. Standard SS1 LSTTL chips are used. Switching transistors can be used in place of the solid-state relays to improve the circuit's efficiency.

COMPRESSOR PROTECTOR



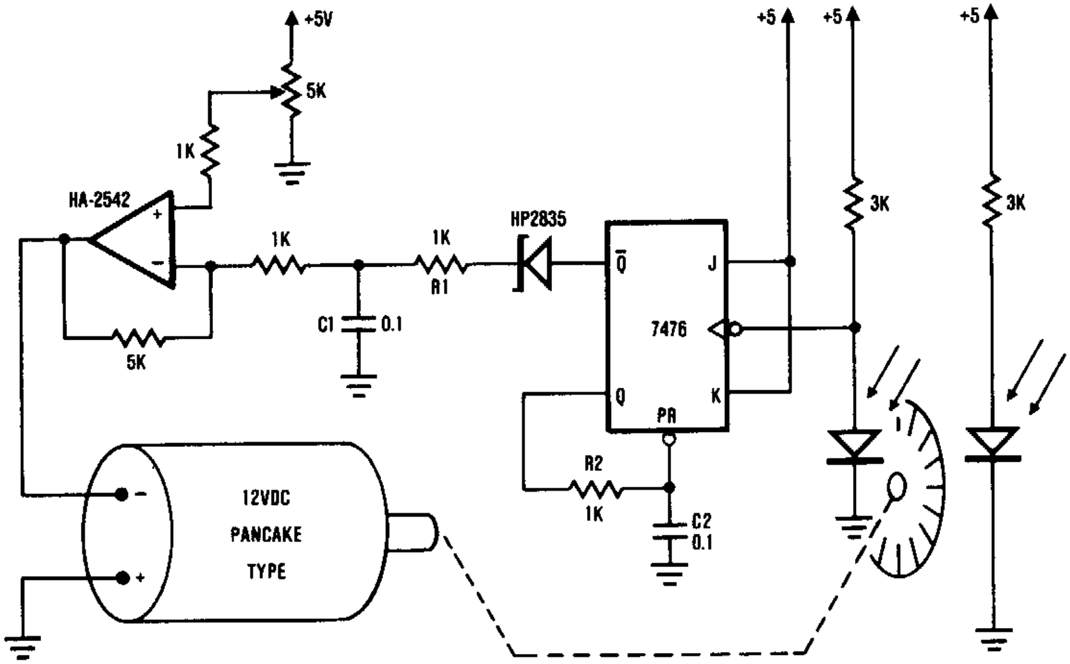
POPULAR ELECTRONICS

Fig. 56-6

This circuit monitors the power-line (ac) voltage. When a power failure occurs, on restoration of power, the circuit adds a five-minute delay before energizing K1, which protects the compressor against limited low voltage.

U1 is a 16-stage counter with an integral oscillator that is set to divide by 8192. R7, R8, and C4 set the oscillator frequency to about 25 Hz, which produces a total count interval of 300 seconds (5 minutes). After this time, pin 8 U1 goes high, which forward biases Q1, triggers SCR1, and activates K1. Up to 30 A can be switched.

dc MOTOR-SPEED CONTROL

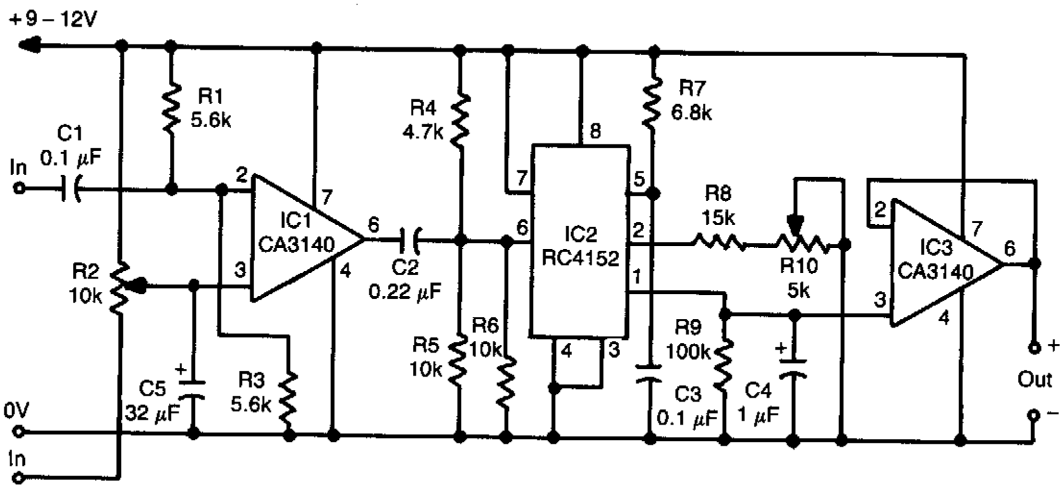


HARRIS

Fig. 56-7

The system shown consists of the HA-2542, a small 12-Vdc motor, and a position encoder. During operation, the encoder causes a series of "constant-width" pulses to charge C1. The integrated pulses develop a reference voltage, which is proportional to motor speed and is applied to the inverting input of HA-2542. The noninverting input is held at a constant voltage, which represents the desired motor speed. A difference between these two inputs will send a corrected drive signal to the motor, which completes the speed control system loop.

CASSETTE MOTOR SPEED CALIBRATOR



RADIO-ELECTRONICS

Fig. 56-8

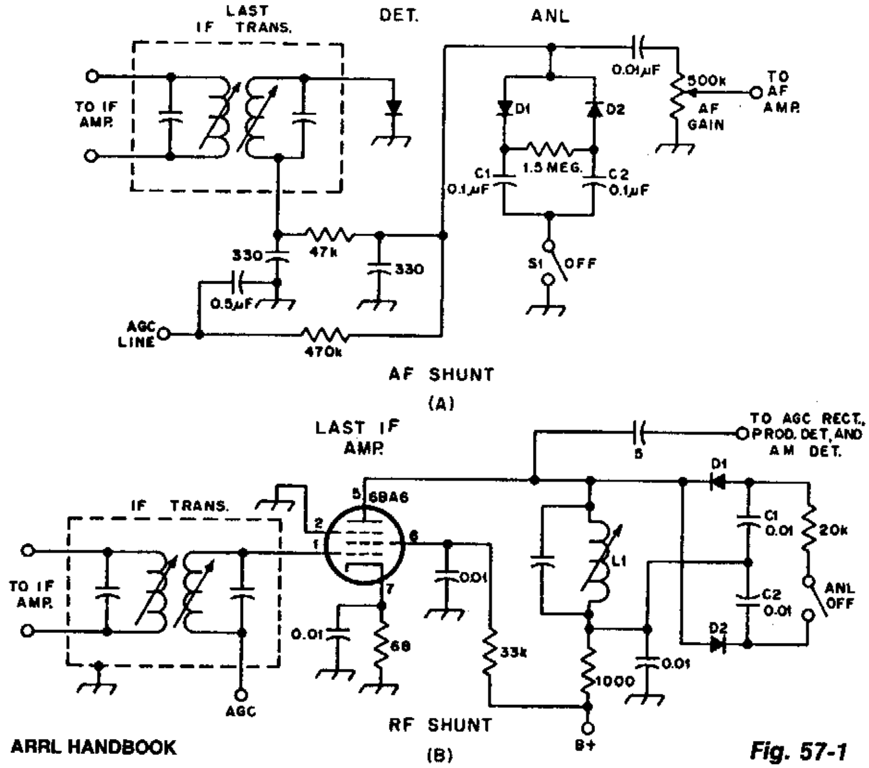
This frequency/voltage converter enables calibrations of cassette-deck speed. It records a steady 1-kHz tone on the cassette deck, and monitors the frequency on the converter. Then, play the tone back and adjust the motor speed until you get the same frequency reading on the converter. The frequency/voltage converter can drive an analog meter to indicate small variations in tape speed.

57

Noise-Reduction Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 672. The figure number in the box of each circuit correlates to the entry in the Sources section.

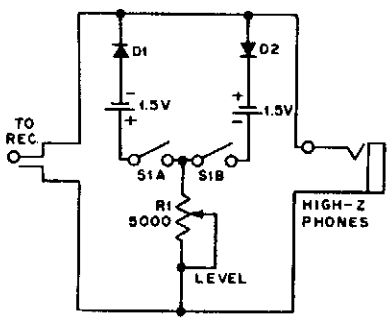
AUDIO SHUNT NOISE LIMITER



ARRL HANDBOOK *Fig. 57-1*

Examples of RF and audio ANL circuits. Positive and negative clipping occurs in both circuits. The circuit at A is self-adjusting. This noise limiter operates at the IF output. It is self-adjusting. Adequate gain is needed at the IF frequency so that several volts p-p of audio is available.

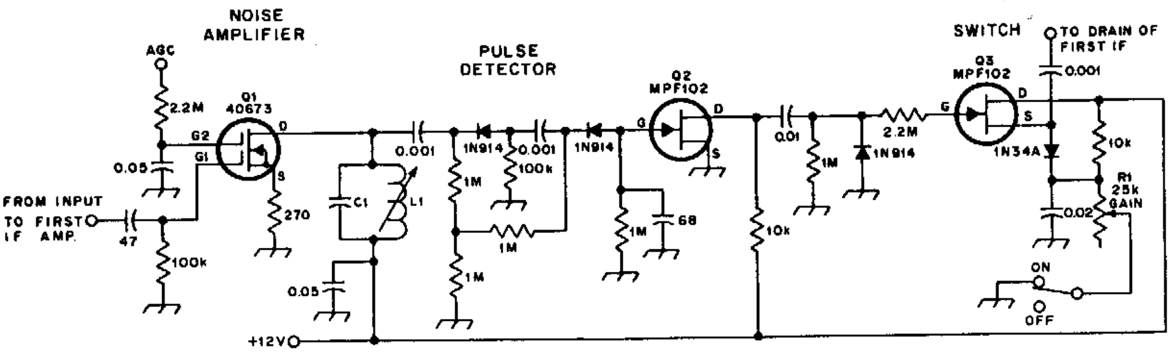
SIMPLE AUDIO CLIPPER/LIMITER



ARRL HANDBOOK *Fig. 57-2*

For use with headphones, this circuit sets the audio clipping level via a 5-kΩ pot. This type of noise clipper works best for pulse-type noise of low duty cycle, such as ignition noise. R1 sets the bias on the diodes for the desired limiting level.

NOISE BLANKER



ARRL HANDBOOK

Fig. 57-3

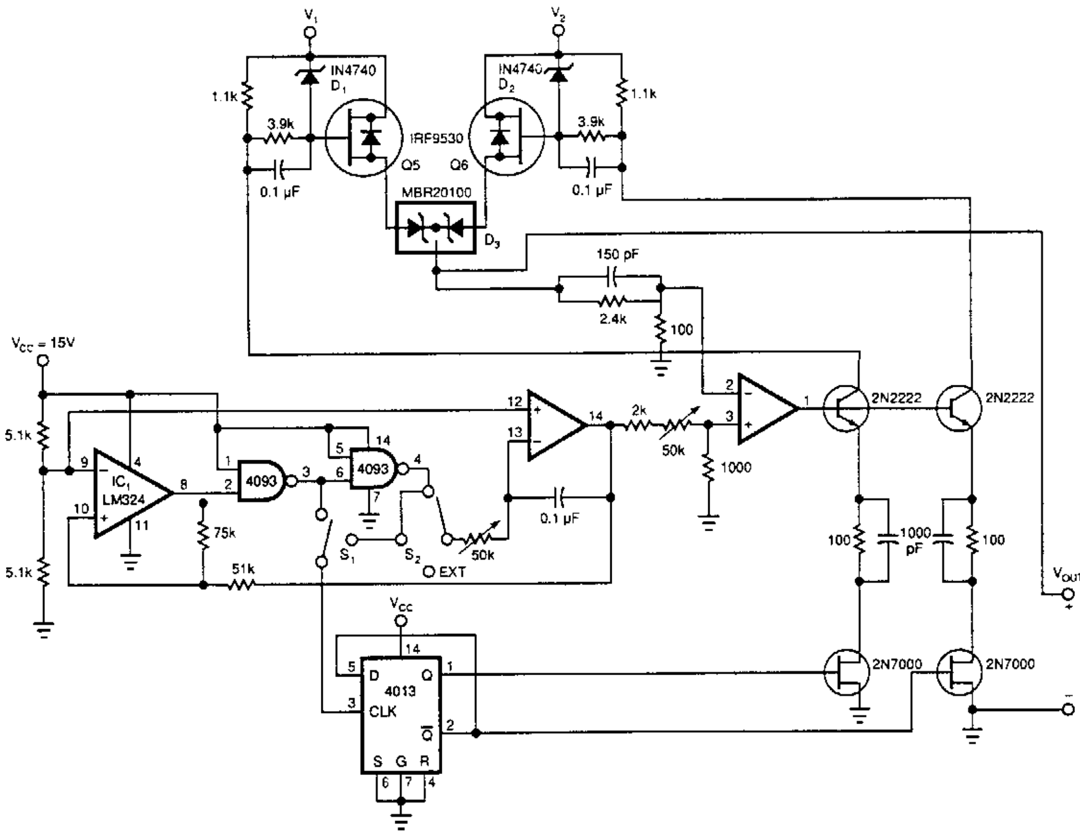
This noise blanker takes a sample of IF input voltage, amplifies it, and drives a switch. The switch (when activated by a noise pulse) adds a heavy load to the first IF stage and kills the gain for the pulse duration.

58

Operational-Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 672. The figure number in the box of each circuit correlates to the entry in the Sources section.

EFFICIENT POWER BOOSTER



EDN

Fig. 58-1

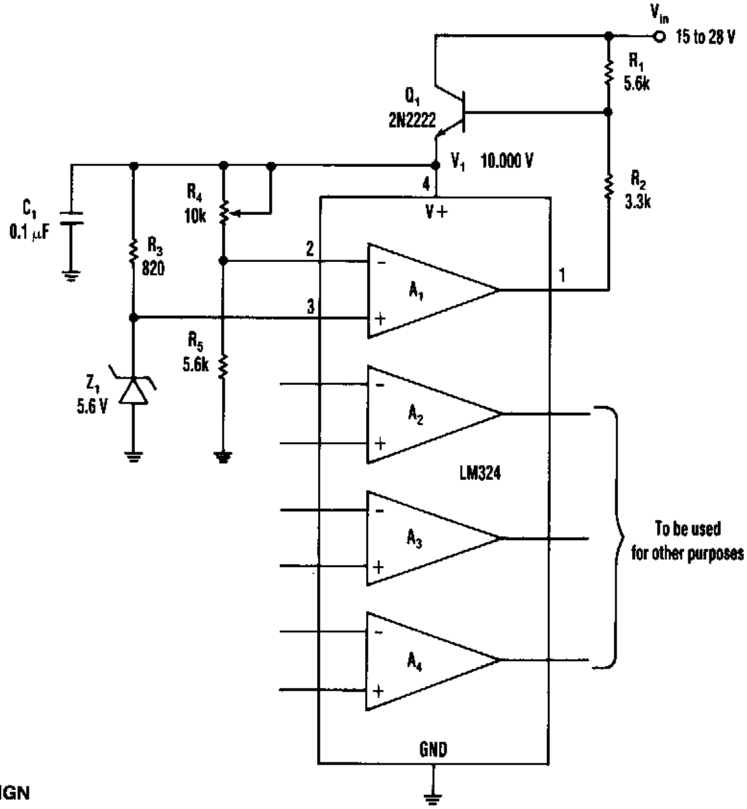
This power booster functions as a high-efficiency "power multiplexer" or, if you supply an external signal-source, as a high-power linear amplifier.

If you want to drive a load with a high-power square wave, the circuit simply draws power from two external power sources, V1 and V2, alternately. In this mode, the circuit's power-handling devices function as switches, dissipating minimal power. The RC time constant of the integrator, IC1, determines the circuit's oscillation period.

If you supply an external drive waveform, the circuit functions as a linear amplifier, and, consequently, inherently dissipates varying portions of that power. The power amplifier is stable for gains ≥ 15.

Diodes D1 and D2 limit the FET's gate-voltage swing to less than 15 V. D3 is a dual Schottky diode that protects the FETs from short circuits between the two supplies, V1 and V2, through a FET's parasitic diode. With D3 in place, you can choose either power channel for the higher voltage input. To drive the FETs, Q5 and Q6, at switching frequencies greater than 1 kHz, you will have to use gate drivers for them.

OP AMP REGULATOR



ELECTRONIC DESIGN

Fig. 58-2

This op amp offers a straightforward method of developing a single-polarity stable voltage source (see the figure). Transistor Q1 gets a base drive through resistor R1, and conducts to develop a voltage (V_1) across the IC's supply pins. Amp A1, R2, and Q1 form a positive-feedback closed loop, along with R3 and the zener diode. A1, R2, and Q1 also form a negative-feedback closed loop with R4 and R5.

The effect of positive feedback is predominant as the noninverting input receives V_1 while the inverting input receives only:

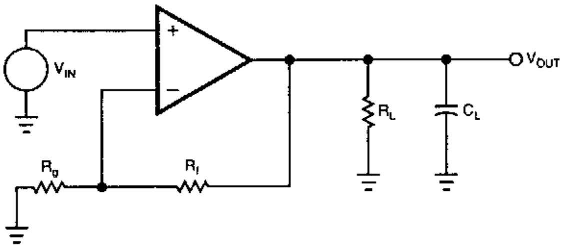
$$V_1 \times \frac{R_5}{R_4 + R_5}$$

This happens until the zener comes into play. When the voltage at the inverting input exceeds the voltage at the noninverting input, A1's output takes away Q1's base current through R2, which reduces V_1. Hence, an equilibrium condition is reached. Now:

$$V_1 = \frac{V_Z(R_4 + R_5)}{R_5}$$

This circuit can source more than 30 mA.

INCREASED FEEDBACK-STABILIZED AMPLIFIER



EDN

Fig. 58-3(a)

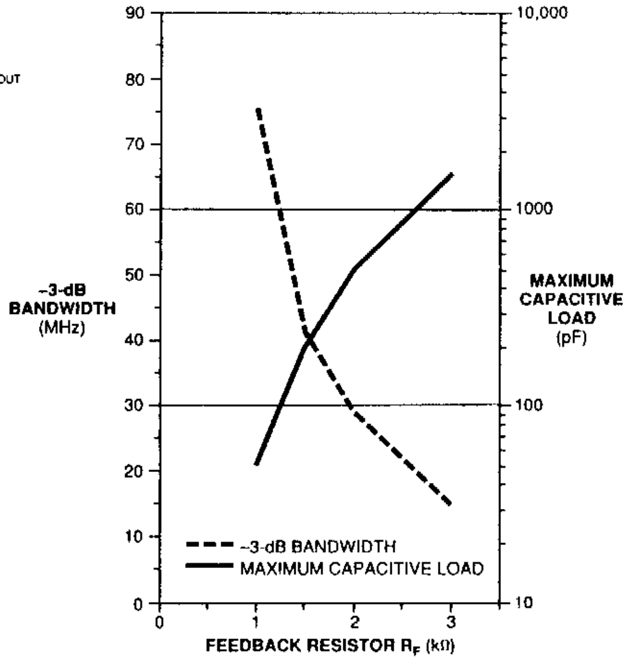


Fig. 58-3(b)

The usual method for using a current-feedback amplifier to drive a capacitive load isolates the load with a resistor in series with the amplifier's output.

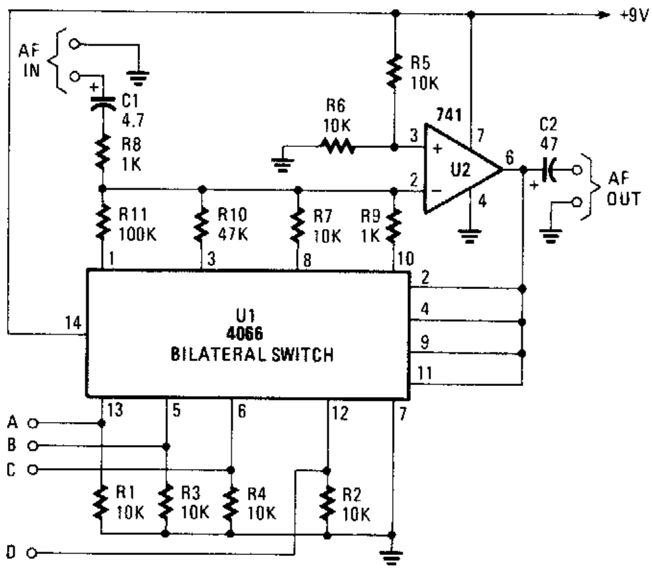
A better solution involves only the amplifier's feedback resistors (Fig. 58-3(a)). Because the feedback resistors determine the amplifier's compensation, you can select the optimal value for these feedback resistors for almost any capacitive load.

Feedback resistance R_F sets the amplifier's bandwidth. Increasing R_F reduces the amplifier's bandwidth, which significantly improves the amplifier's ability to drive capacitive loads. Feedback resistor RG sets the amplifier's gain.

You cannot get the data necessary to calculate alternate values for R_F from most data sheets. However, a few minutes at the bench with a network analyzer will generate the data to make a graph of the value of the feedback resistor vs. the amount of capacitive load the amplifier can drive (Fig. 58-3(b)).

Start with the recommended data-sheet value for feedback resistor R_F and measure the amplifier's frequency response without any capacitive load. Note the bandwidth, then add capacitive loading until the response peaks by about 5 dB. Record this value of capacitance; it is the maximum amount for that feedback resistor. Then, increase the value of the feedback resistor and repeat the procedure until you develop a graph like the one in Fig. 58-3(b).

GAIN-CONTROLLED OP AMP

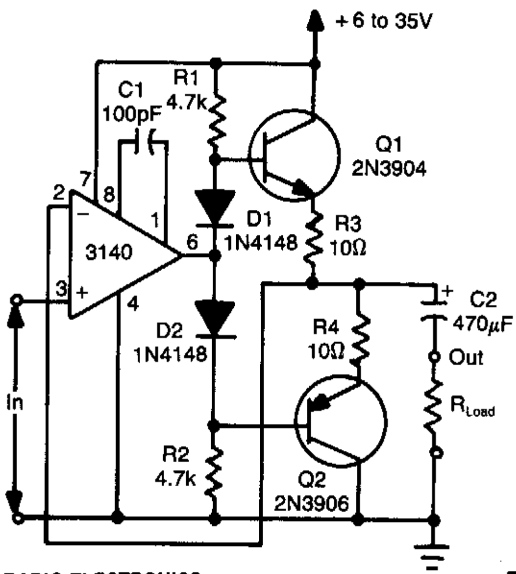


POPULAR ELECTRONICS

Fig. 58-4

The gain controller uses a 4066 quad bilateral switch to electronically select a feedback resistor for the 741 op amp. One or more switches can be turned on at the same time to produce a stepped, variable-gain range from less than 1 to 100.

BIDIRECTIONAL COMPOUND OP AMP

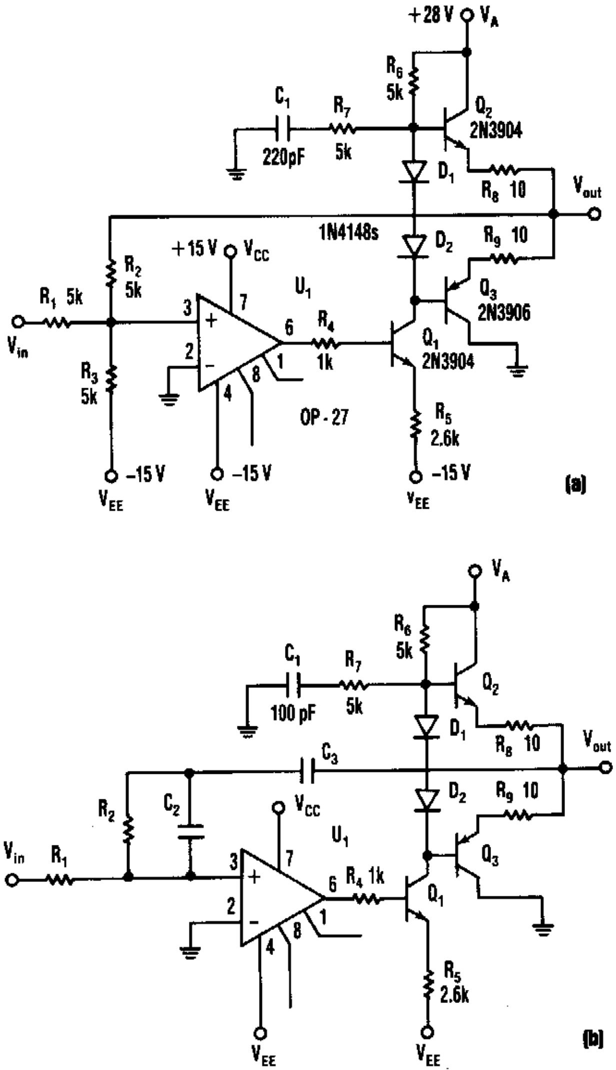


RADIO-ELECTRONICS

Fig. 58-5

Using two transistors (Q1 and Q2), a bidirectional op amp can source or sink up to 50 mA. D1 and D2 provide bias for Q1 to eliminate "dead-zone" effects.

COMPOUND OP AMP VCO DRIVER

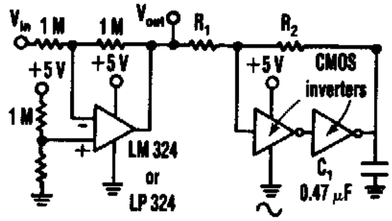


ELECTRONIC DESIGN

Fig. 58-6

This circuit produces 5- to 25-V output to drive a VCO from a standard ±15-V supply system. R7 and C1 supply frequency compensation. Q1 through Q3 form an inverting amplifier with a gain of two. Negative feedback through R2 closes the loop. This circuit can act as a an active load-log filter and directly drive a voltage-controlled oscillator.

MAKE LM324 OP AMP SWING RAIL-TO-RAIL

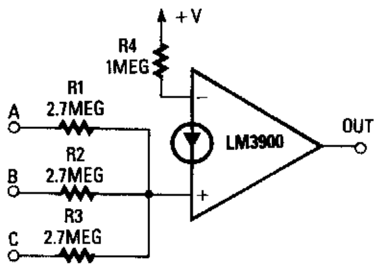


	TYPICAL OUTPUT SWING					
Op Amp	R_1 =	Open circuit	2.2 k	1 k	330 Ω	150 Ω
	R_2 =	Don't care	15 k	4.7 k	1.5 k	470 Ω
LM324	V_{out} Hi =	3.71 V	4.30 V	4.73 V	4.89 V	4.94 V
	V_{out} Lo =	0.04 V	0.023 V	0.015 V	0.010 V	0.003 V
LP324	V_{out} Hi =	4.16 V	4.91 V	4.965 V	4.982 V	4.987 V
	V_{out} Lo =	0.53 V	0.064 V	0.022 V	0.007 V	0.003 V

ELECTRONIC DESIGN *Fig. 58-7*

By using two CMOS inverters, the output for an LM324 op amp can be increased from 3.5 Vpp to 4.9 Vpp. This circuit is only recommended for light loads (<30 mA) and for relatively slow op amps. Any CMOS inverter (74COO, 74CO2, 74C14, CD4001, CD4011, etc.) can be used.

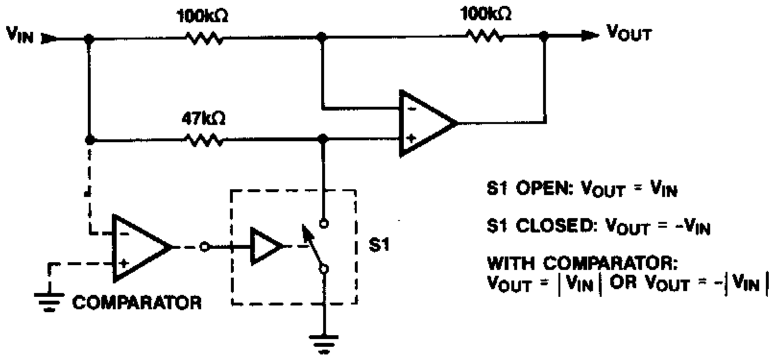
3-INPUT AND GATE COMPARATOR



RADIO-ELECTRONICS *Fig. 58-8*

This circuit has high output only when all three inputs are high. The noninverting-input current, when all three inputs are high, must exceed that of the inverting input, as determined by R4. The circuit can be converted to a NAND gate by transposing the two inputs of the op amp.

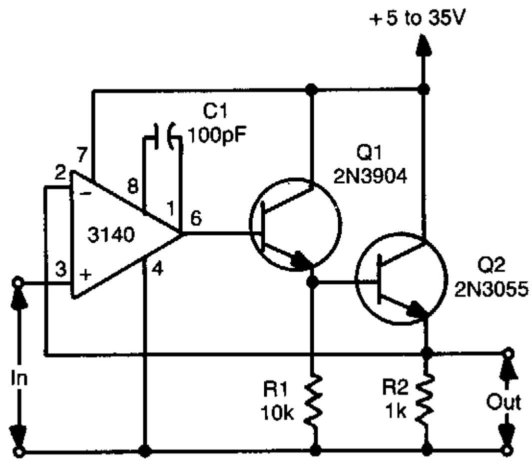
PROGRAMMABLE INVERTER/RECTIFIER



MAXIM *Fig. 58-9*

The op amp is alternately an inverter or buffer, under control of the switch polarity. As a buffer, the gain is always 1, but as an inverter, the gain is set by the ratio of the input and feedback resistors. By adding a comparator, the function can be synchronously switched as the input polarity changes, which effectively rectifies the output. The output polarity is determined by the switch logic (normally open or normally closed) and the comparator input polarity.

COMPOUND OP AMP



RADIO-ELECTRONICS *Fig. 58-10*

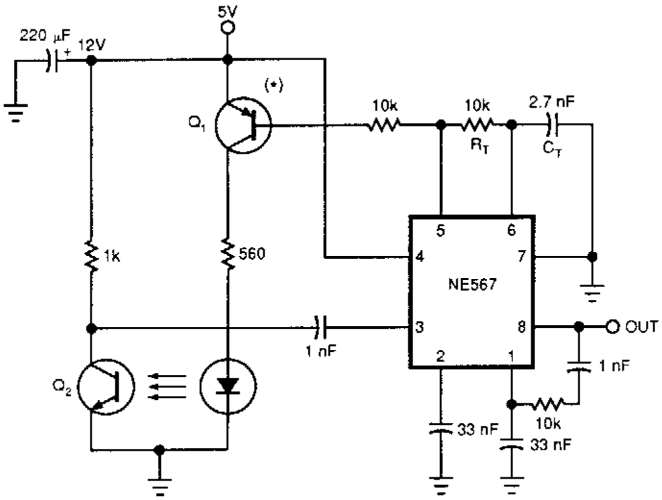
By using an emitter-follower or a Darlington pair, a voltage-follower op amp configuration can source higher currents than the op amp otherwise could.

59

Optical Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 672. The figure number in the box of each circuit correlates to the entry in the Sources section.

OPTICAL INTERRUPTION SENSOR



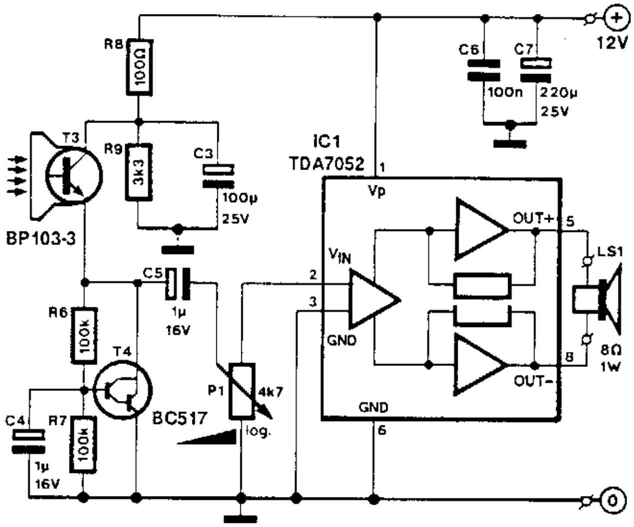
Source	Detector	Manufacturer	Notes
TIL 32	TIL 78	Texas Instruments	IR LED and phototransistor
TIL 38	TIL 414	Texas Instruments	IR LED and phototransistor
CQY58A	BPW22A	Philips	IR LED and phototransistor
CQY89A	BPW 50	Philips	IR LED and pin photodiode
TIL 139		Texas Instruments	Transmissive source and detector assembly
TIL 149		Texas Instruments	Reflective source and detector assembly

EDN ***Fig. 59-1***

Using only an 8-pin IC and a few discrete components, you can build the infrared optical interrupter. The NE567 tone decoder has all the necessary circuit elements: a local oscillator, a PLL decoder, and a 100-mA output-drive capability. The local oscillator, which is tuned to 40 kHz by RT and CT, drives Q1, a universal low-power silicon pnp transistor (such as a 2N3906, BC559, or ZTX500). Q1 drives the IR-emitting diode. The receiving part of the circuit surrounds the IC's internal PLL input at pin 3. When the photodetector, Q2, detects the oscillating IR light beam, the 40-kHz signal appears at pin 3 of the IC. Under this condition, the circuit locks and the IC's output is high. When something opaque comes between the LED and Q2, the 40-kHz signal doesn't reach the PLL input, and the IC's output goes low.

The feedback network between pins 1 and 8 prevents the output from chattering. If you connect this circuit to a high-inertia load (such as a mechanical relay), the output doesn't tend to oscillate and you can eliminate these feedback components. The circuit works with virtually any LED-photodetector pair, but matched pairs allow for longer distances between the emitter and receiver. The table lists some of the best choices.

LIGHT RECEIVER

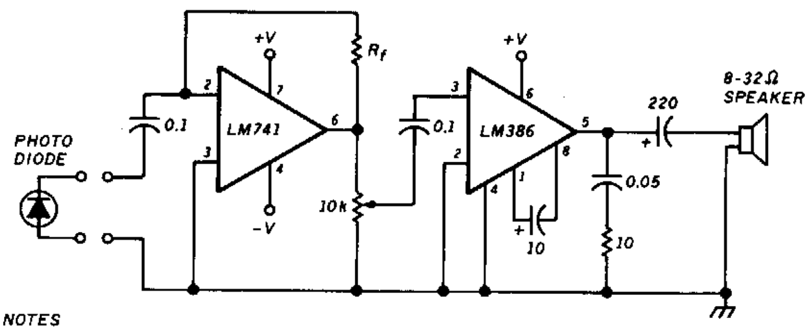


ELEKTOR ELECTRONICS

Fig. 59-2

T3 is a photocell or phototransistor. T4 controls the emitter voltage of T3. IC1 is an audio amplifier to provide amplification of the signal from the photocell.

OPTICAL RECEIVER



NOTES

+V = +9V TO +12V

-V = -9V TO -12V

R_f = 100k - 1M THIS RESISTOR DETERMINES THE GAIN OF THE FIRST AMPLIFIER STAGE

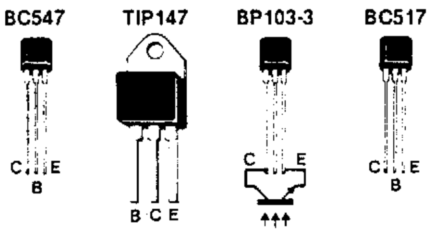
A LOW NOISE OP-AMP MAY BE SUBSTITUTED FOR THE LM741 FOR IMPROVED PERFORMANCE

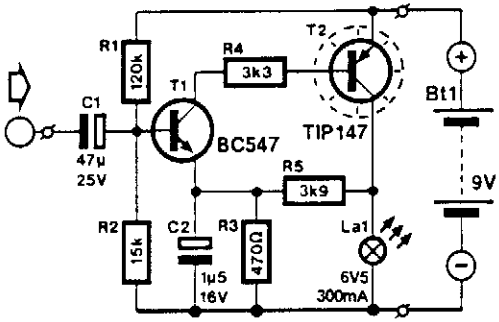
HAM RADIO

Fig. 59-3

An optical receiver for light-wave communications, this circuit works with AM-type light signals.

LIGHT TRANSMITTER

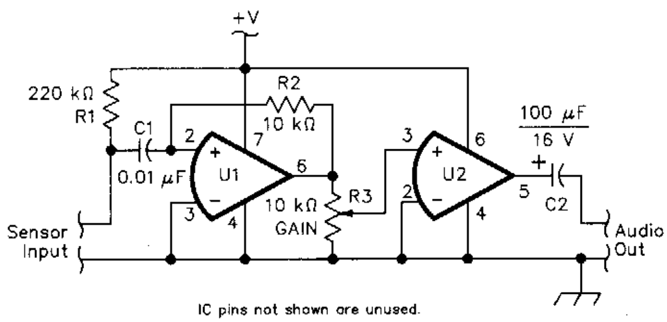


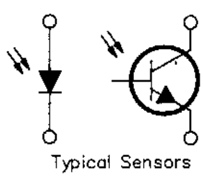


ELEKTOR ELECTRONICS *Fig. 59-4*

This circuit modulates the current through a lamp filament. Use a low-voltage lamp with a thin, straight filament. They have a fast response to filament voltage variations. dc is applied to "bias" the filament that is on, and the audio is superimposed. A BC457 drives a TIP147, which modulates the filament current.

OPTICAL/LASER RECEIVER

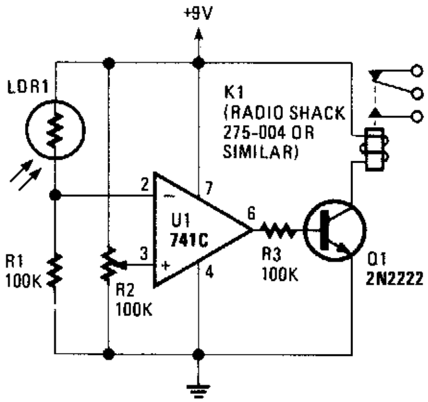




QST *Fig. 59-5*

Using a single 741 op amp, a photodiode sensor, and an LM386, this simple receiver operates from a 9-V battery. The circuit will drive a pair of earphones or a small speaker.

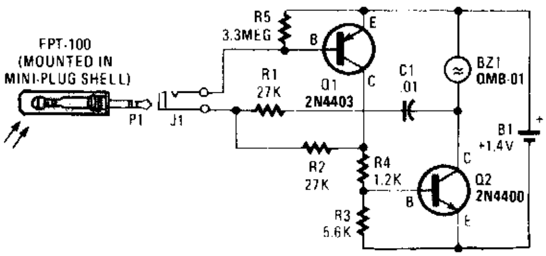
LIGHT DETECTOR



HANDS-ON ELECTRONICS *Fig. 59-6*

The circuit's threshold is set by resistor R2. When the intensity of the light falling on the LDR is lowered, resistance of that unit increases, and lowers the voltage applied to the inverting input of the 741. The reference voltage at the noninverting input of the 741 is set (via R2) so that the comparator switches from low to high when the light falling on the LDR is reduced. That high activates transistor Q1, which causes the relay contacts to close.

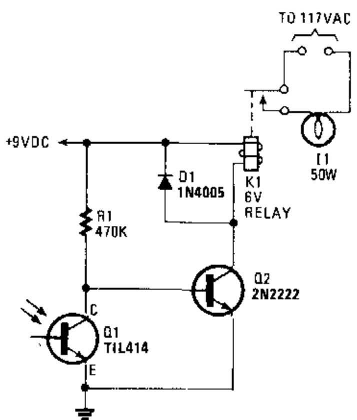
LIGHT PROBE



POPULAR ELECTRONICS *Fig. 59-7*

Originally designed as an aid for blind people, this probe was used as a light detector in order to tell if a device or room lights are on or off.

SIMPLE PHOTOELECTRIC LIGHT CONTROLLER



POPULAR ELECTRONICS *Fig. 59-8*

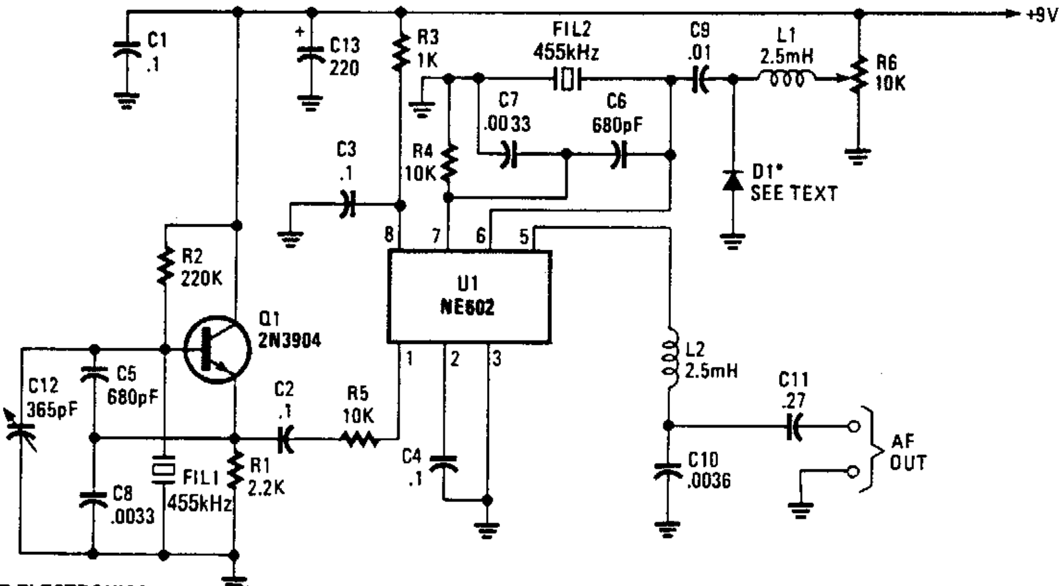
A phototransistor senses daylight. At dusk, it ceases to conduct and R1 biases Q2, activates K1, and switches on the light. At dawn, Q1 starts to conduct, and Q2 is cut off. K1 drops out and the light goes out.

60

Oscillators

The sources of the following circuits are contained in the Sources section, which begins on page 672. The figure number in the box of each circuit correlates to the entry in the Sources section.

BEAT-FREQUENCY AUDIO GENERATOR

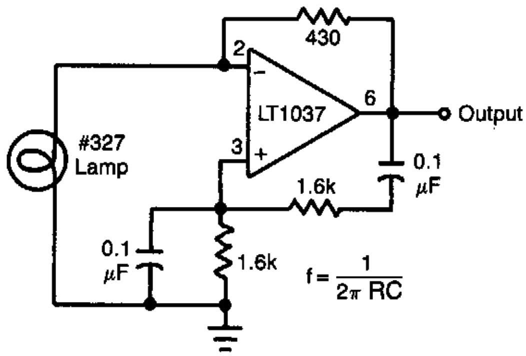


POPULAR ELECTRONICS

Fig. 60-1

Q1 is a fixed oscillator operating at 455 kHz. U1 is a mixer, with its own internal oscillator running at 455 kHz. FIL1 and FIL2 are Murata CSB455E filters or equivalent. D1 is a varactor diode (an IN4002 used as a varactor works well here). R6 controls the bias on D1. When R6 is varied, the oscillator frequency varies a few kHz. Audio beat note is taken, through RF filter L2 and C10, from pin 5 of U1.

SIMPLE WIEN-BRIDGE OSCILLATOR

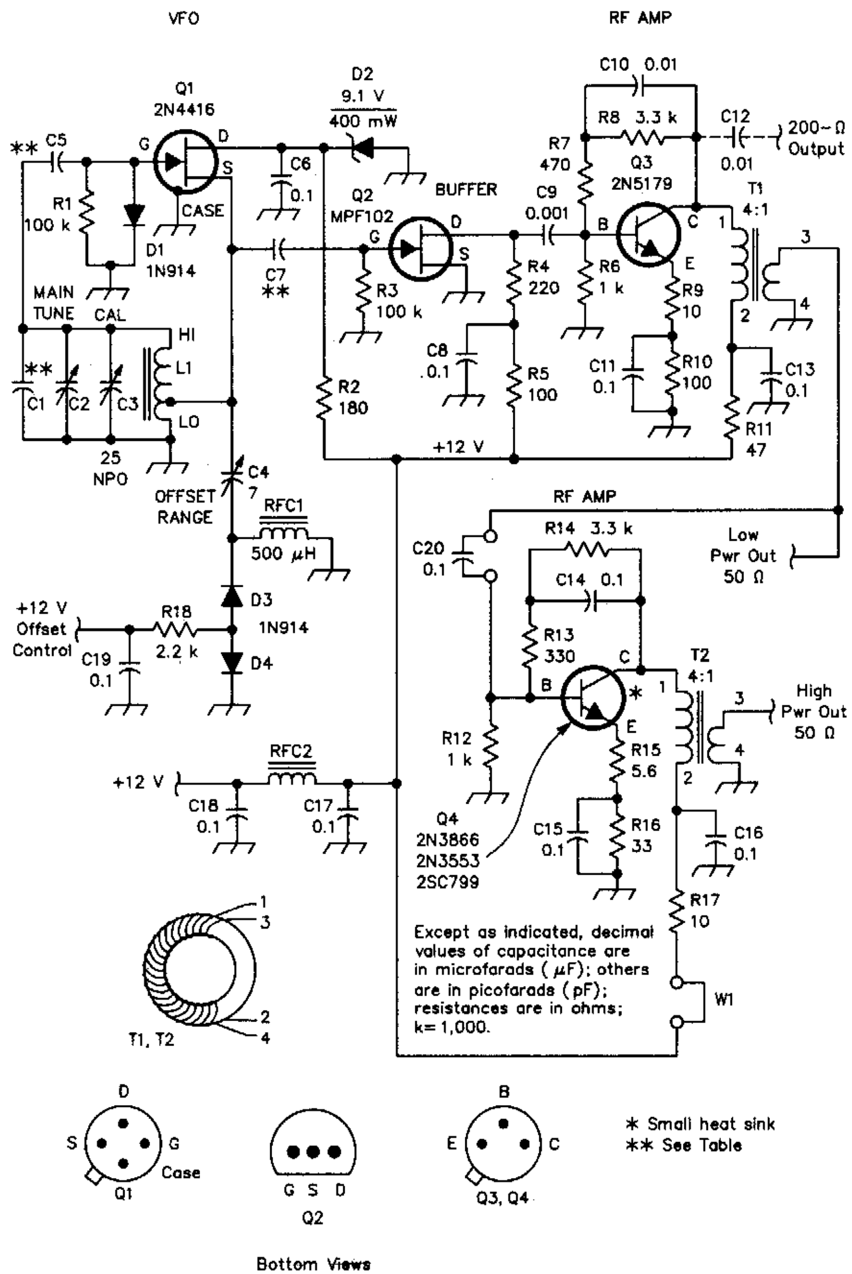


EDN

Output Frequency = 1,000 kHz
For Values Given

Fig. 60-2

In this circuit, the Wien-bridge network provides phase shift, and the lamp regulates the amplitude of the oscillations. The smooth, limiting nature of the lamp's operation, in combination with its simplicity, gives good results. Harmonic distortion is below 0.3%.


VFO
RF AMP
Q1
2N4416
D2
9.1 V
400 mW
C10 0.01
R8 3.3 k
C12
200-Ω
Output
0.01
R7
470
Q3
2N5179
T1
4:1
** C5
G
D
S
C6
0.1
CASE
Q2
MPF102
BUFFER
C9
0.001
R1
100 k
D1
1N914
C7
**
R3
100 k
R4
220
R6
1 k
R9
10
MAIN
TUNE
CAL
HI
L1
LO
**
C1
C2
C3
25
NPO
C8
.0.1
R5
100
C11
0.1
R10
100
C13
0.1
R2
180
+12 V
R11
47
C4
7
OFFSET
RANGE
RFC1
500 µH
RF AMP
C20
0.1
R14 3.3 k
C14 0.1
Low
Pwr Out
50 Ω
D3
1N914
+12 V
Offset
Control
R18
2.2 k
C19
0.1
D4
R13
330
T2
4:1
High
Pwr Out
50 Ω
R12
1 k
R15
5.6
+12 V
RFC2
C18
0.1
C17
0.1
Q4
2N3866
2N3553
2SC799
C15
0.1
R16
33
C16
0.1
R17
10
W1
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k= 1,000.
T1, T2
Case
Q1
G S D
Q2
Q3, Q4
* Small heat sink
** See Table
Bottom Views

QST

Fig. 60-3

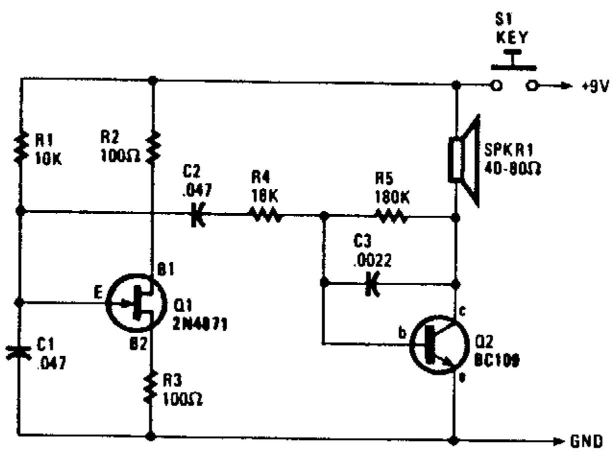
This VFO circuit covers from 2.13 to 2.58 MHz and is intended for use with an external mixer to heterodyne the signal to desired frequencies. Coil data is shown in the parts list. Two power output levels are available—a few hundred mW (0 to +3 dbm) from Q3. Q4 is a class A amplifier for boosting the power to +22 dbm for driving a high-level mixer. The VFO can be operated on other frequencies, with suitable component charges (see the table).

Component Information

f(MHz)	*C1(pF)*	*C2(pF)*	*C5, C7(pF)*	*L1*
1.8-2	220	100	100	24 μH; 71 turns of no. 30 enamel on a T-68-6 toroid core. Tap at 18 turns from bottom end.
3.5-4	150	50	68	9.5 μH; 44 turns of no. 26 enamel on a T-68-6 toroid core. Tap at 11 turns from bottom end.
5-5.5	130	50	47	5 μH; 33 turns of no. 24 enamel on a T-68-6 toroid core. Tap at 8 turns from bottom end.
7-7.3	110	25	47	3.6 μH; 27 turns of no. 24 enamel on a T-68-6 toroid core. Tap at 7 turns from bottom end.

C3 is a 25-pF NPO ceramic trimmer. C4 is a 7-pF air trimmer. The total capacitance of C2 may be reduced to restrict the VFO tuning range. C1, C5 and C7 are NPO ceramic.

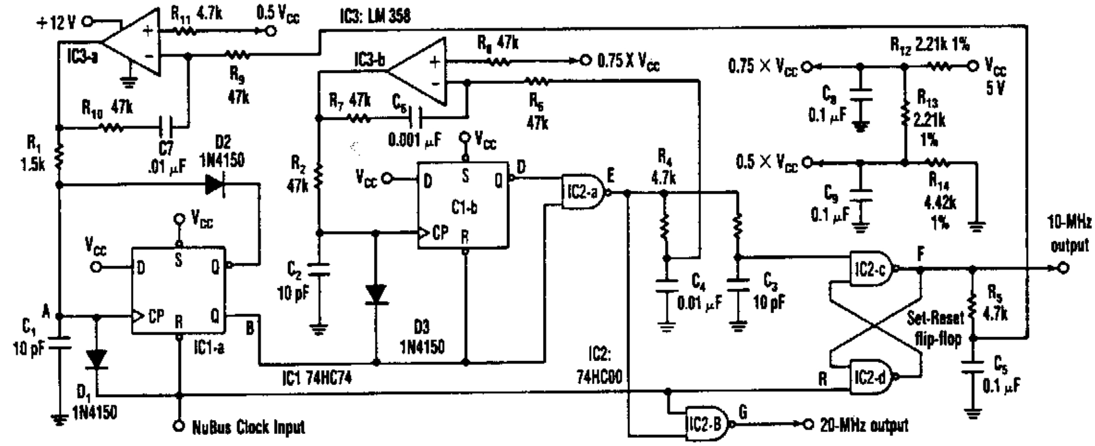
CODE-PRACTICE OSCILLATOR I



WELS' THINK TANK *Fig. 60-4*

Q1, a unijunction transistor, generates a sawtooth of about 1.5 to 2 kHz, depending on C1 and R1. Q2 acts as a speaker driver. A 9-V battery is used, and the keying is done by keying the supply line.

PHASE-LOCKED 20-MHz OSCILLATOR

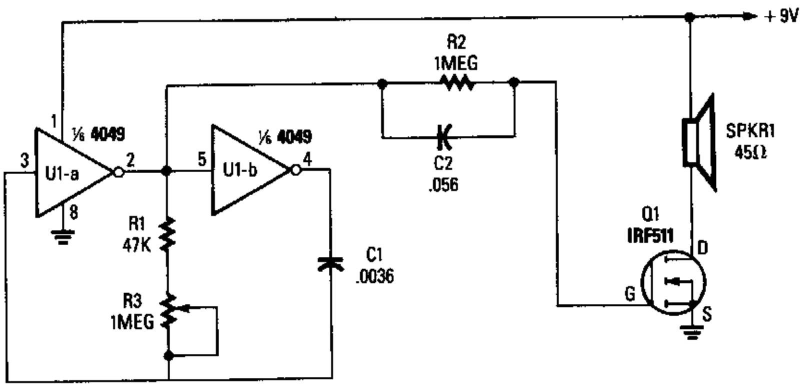


ELECTRONIC DESIGN

Fig. 60-5

This circuit produces a 20-MHz clock phase locked to a 10-MHz clock present in the Apple MAC II. To generate the 20-MHz signal, the circuit produces a 25 ns negative-going pulse delayed 50 ns from the falling edge of the 10-MHz Nubus clock input at point E. NORing that pulse with the Nubus clock produces the 20-MHz clock at point G. Applying the 25-ms pulse to the set input of an S/R flip-flop and the Nubus clock to the reset input results in a 10-MHz square wave at F.

AUDIO OSCILLATOR



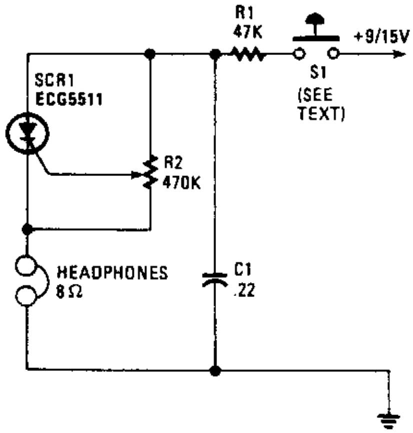
POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

Fig. 60-6

Two gates, U1A and U1B (1/3 of a 4049 hex inveter), are connected in a VFO circuit. Components R1, R3, and C1 set the frequency range of the VFO. With the values given, the circuit's output can range from a few hundred hertz to over several thousand hertz by adjusting R3.

The simplest way to change the frequency range of the oscillator is to use different capacitance values for C1. A rotary switch, teamed up with a number of capacitors, can be used to select the desired frequency range.

CODE-PRACTICE OSCILLATOR II

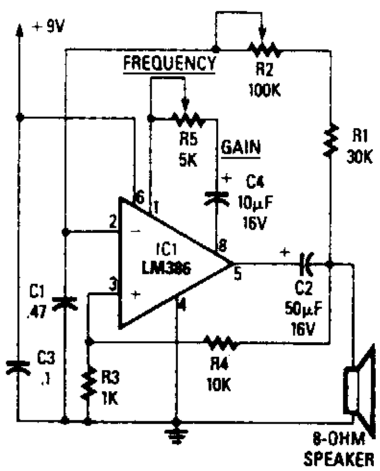


Capacitor C1 charges through resistor R1, and when the gate level established by potentiometer R2 is high enough, the SCR is triggered. Current flows through the SCR and earphones, discharging C1. The anode voltage and current drop to a low level, so the SCR stops conducting and the cycle is repeated. Resistor R2 lets the gate potential across C1 be adjusted, which charges the frequency or tone. Use a pair of 8-Ω headphones. The telegraph key goes right into the B+ line, 9-V battery.

HANDS-ON ELECTRONICS/POPULAR ELECTRONICS

Fig. 60-7

AUDIO OSCILLATOR II



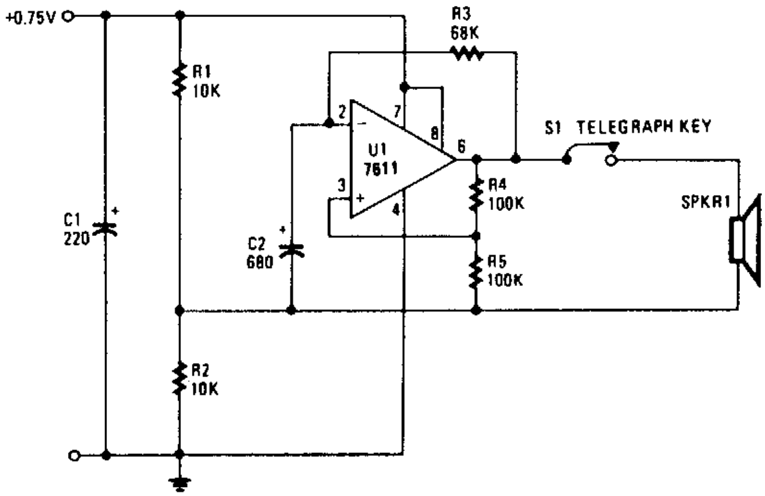
The circuit's frequency oscillation is $f = 2.8/[C_1 \times (R_1 + R_2)]$. Using the values shown, the output frequency can be varied from 60 Hz to 20 kHz by rotating potentiometer R2.

A portion of IC1's output voltage is fed to its noninverting input at pin 3. The voltage serves as a reference for capacitor C1, which is connected to the noninverting input at pin 2 of the IC. That capacitor continually charges and discharges around the reference voltage, and the result is a square-wave output. Capacitor C2 decouples the output.

WELS' THINK TANK

Fig. 60-8

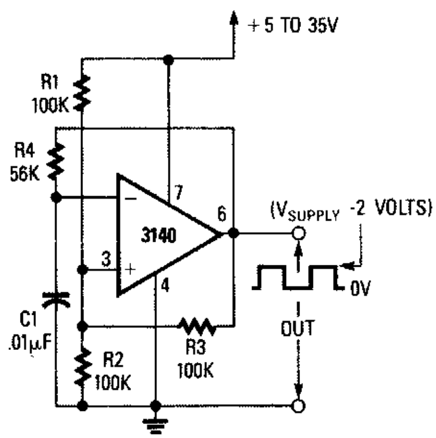
CODE-PRACTICE OSCILLATOR III



WELS' THINK TANK *Fig. 60-9*

U1 is used as an oscillator; the frequency is determined by C2 and R3. Use an 80 Ω or similar high-impedance speaker, and a 1.5-V battery for a power source.

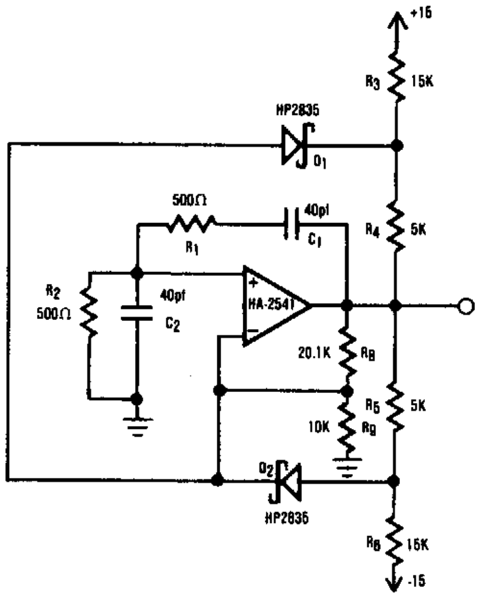
RELAXATION OSCILLATOR



RADIO-ELECTRONICS *Fig. 60-10*

This oscillator runs at about 150 Hz, but C1 and/or R4 can be proportionately changed to alter this frequency. Rise and fall times are 12 and 7 μs, respectively.

WIEN-BRIDGE OSCILLATOR



HARRIS

Fig. 60-11

The HA2541 is well-suited for use as the heart of an oscillator. In spite of the rudimentary diode limiting that is provided by R3 through R7 and D1 and D2, a good-quality sine wave of 40 MHz is readily attainable with an upper limit of 50 MHz, which exceeds the unity-gain bandwidth of HA-2541.

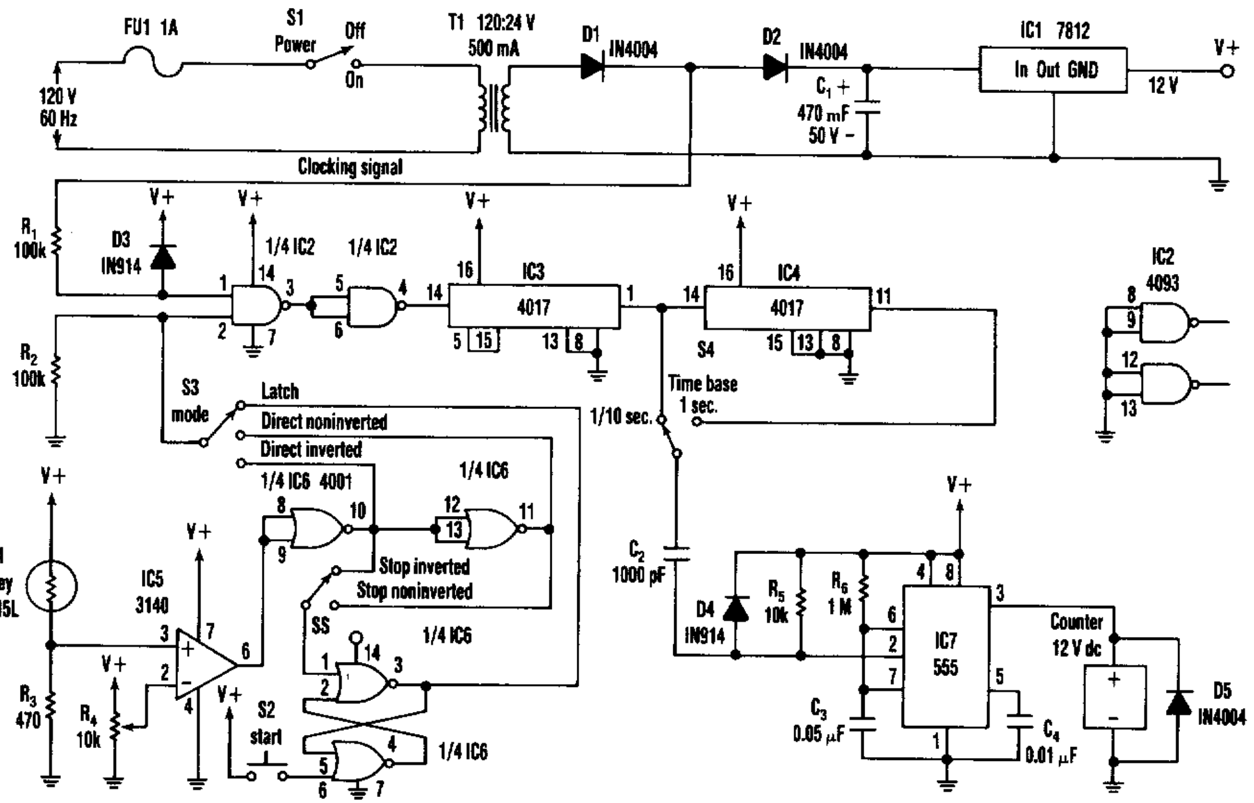
R1/C1 and R2/C2 provide the required regenerative feedback needed for adequate frequency stability. In theory, the feedback network requires a gain of three to sustain oscillation. However, the practical gain needed is just over three and is provided by R8 and R9.

61

Photography-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 673. The figure number in the box of each circuit correlates to the entry in the Sources section.

PHOTO-EVENT TIMER

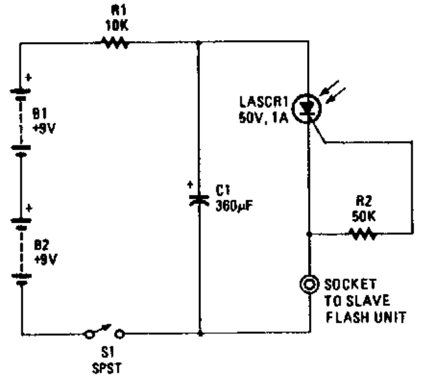


ELECTRONIC DESIGN

Fig. 61-1

S2 is used to initiate timing. A light-to-dark or dark-to-light transition stops this timer, depending on the setting of S5. S3 offers a direct operating mode, rather than through the latch. IC3 and IC4 supply 0.1- or 1-second timing pulses. IC7 drives a time display counter, a 12-Vdc unit that draws less than 200 mA.

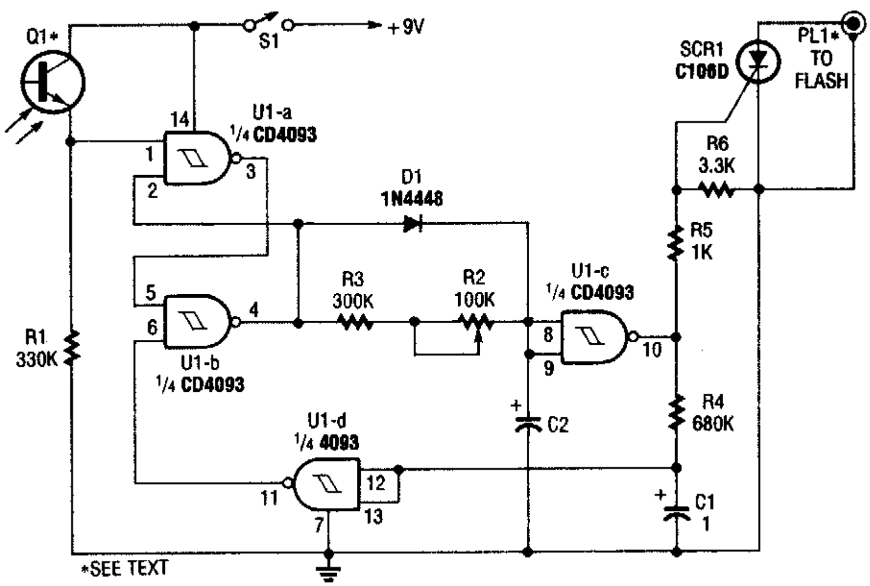
SCR SLAVE FLASH



POPULAR ELECTRONICS *Fig. 61-2*

Using a light-activated SCR, this circuit can trigger a slave-flash unit.

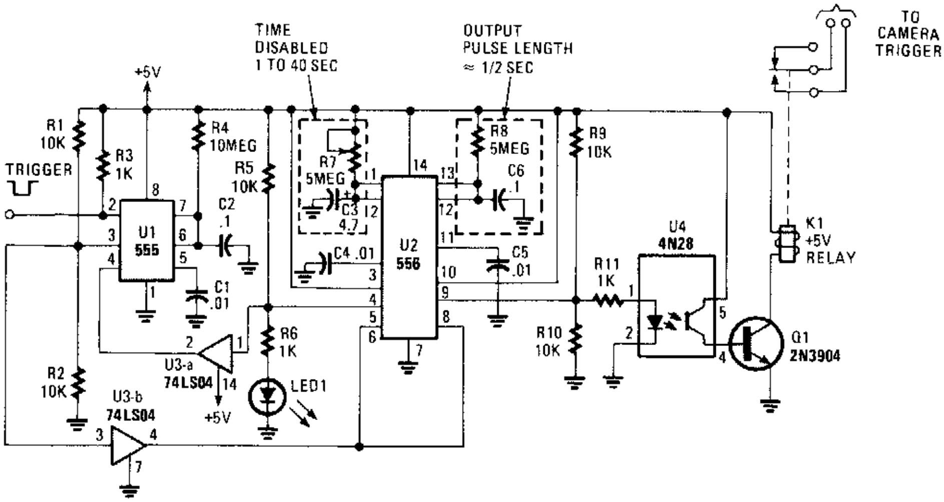
TIME-DELAY PHOTO-FLASH TRIGGER



POPULAR ELECTRONICS *Fig. 61-3*

Q1 is a phototransistor that is normally illuminated by a beam of light. When the beam is intercoupled, pin 1 of U1A goes high, and forces pin 4 U2B low. Then, C2 discharges through R2 and R3. After a certain time delay, pin 10 U1C goes high, triggers SCR1, and sets off the flash. R4/C1 charges, causes U1D output to go low after about $1/2$ second, and resets U1A and U1B to the initial state. This delay prevents accidental double flash exposure.

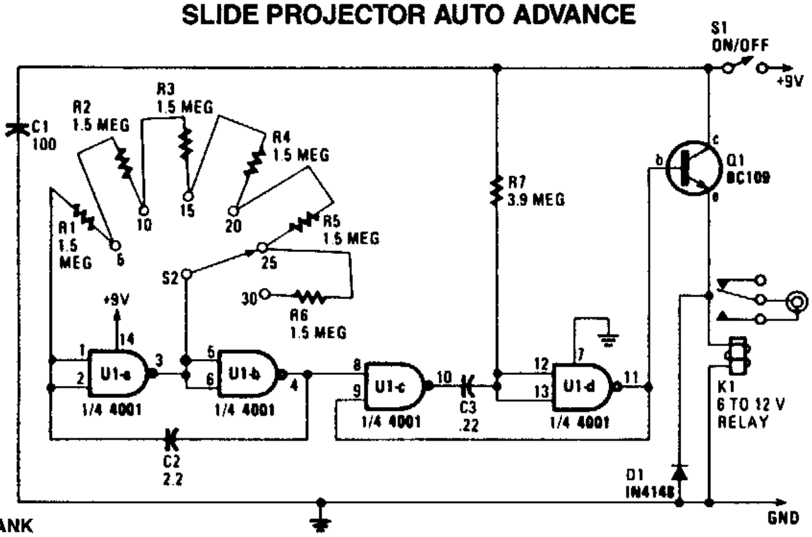
CAMERA TRIP CIRCUIT



POPULAR ELECTRONICS **Fig. 61-4**

This circuit was used to trip a camera shutter. Grounding pin 2 of U1 makes pin 4 of U1 go high. This triggers both timers of dual timer U1. One output holds reset (pin 4) of U1 low to keep U1 from accepting another trigger, depending on the time constant of R7 and C3. This prevents camera film waste. The other timer is used to generate a ½-second pulse to drive U4 and Q1, the relay driver. K1 triggers the camera.

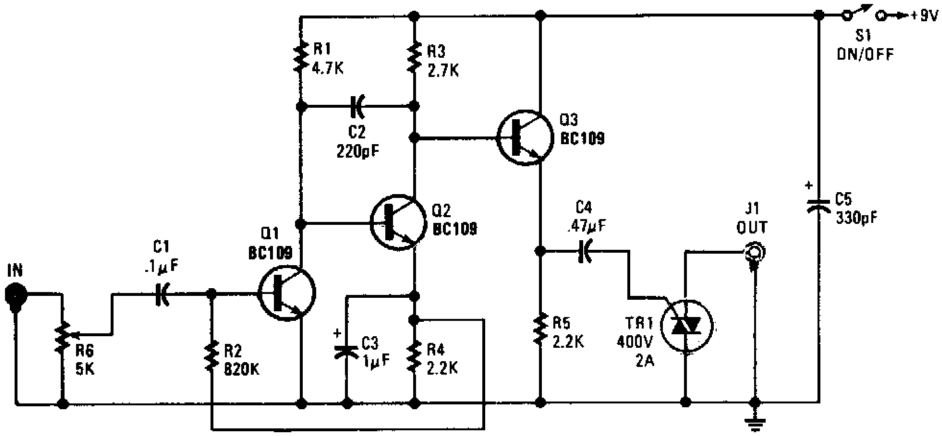
SLIDE PROJECTOR AUTO ADVANCE



WELS' THINK TANK **Fig. 61-5**

A 4001 CMOS Quad NORgate is set up as an astable multivibrator, which drives a simple differentiator and relay driver. Depending on the setting of S2, a delay of 5 to 30 seconds is generated. S2 and R1 through R6 can be replaced by a single 10-MΩ pot, if desired.

SOUND-TRIGGERED FLASH

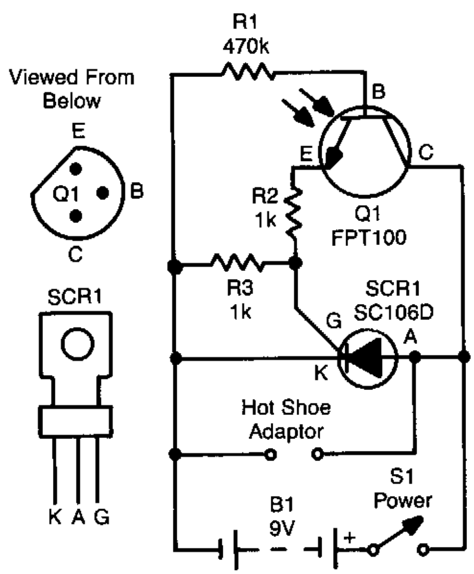


POPULAR ELECTRONICS

Fig. 61-6

Audio input from a microphone drives amplifier Q1/Q2/Q3 to produce an ac voltage across R5. C4 couples this to TR1, causing it to conduct, triggering photoflash or other device that is connected to J1.

SLAVE FLASH TRIGGER



The SCR is wired across the trigger circuit of the flash gun. Normally, the SCR is off, so the flash gun is able to charge to its trigger voltage. Photo transistor Q1 is used to monitor the light level. When a high-intensity flash occurs, Q1 briefly conducts and supplies gate current to the SCR. That causes the SCR to turn on, which then triggers the slave flash gun via the hot-shoe adapter terminals.

Once the flash gun has triggered, the SCR quickly turns off again. That happens because the current in circuit quickly falls below the SCR's holding current. The resistor at the base of Q1 (R1) determines the sensitivity of the circuit. If you wish, you can reduce the sensitivity, simply by reducing the value of the resistor from that shown. The 1-kΩ resistor between the gate and cathode of the SCR (R3) prevents the SCR from false triggering if high voltages are applied between the anode and the cathode. Q1 can also be a GEL14G2.

POPULAR ELECTRONICS/HANDS-ON ELECTRONICS

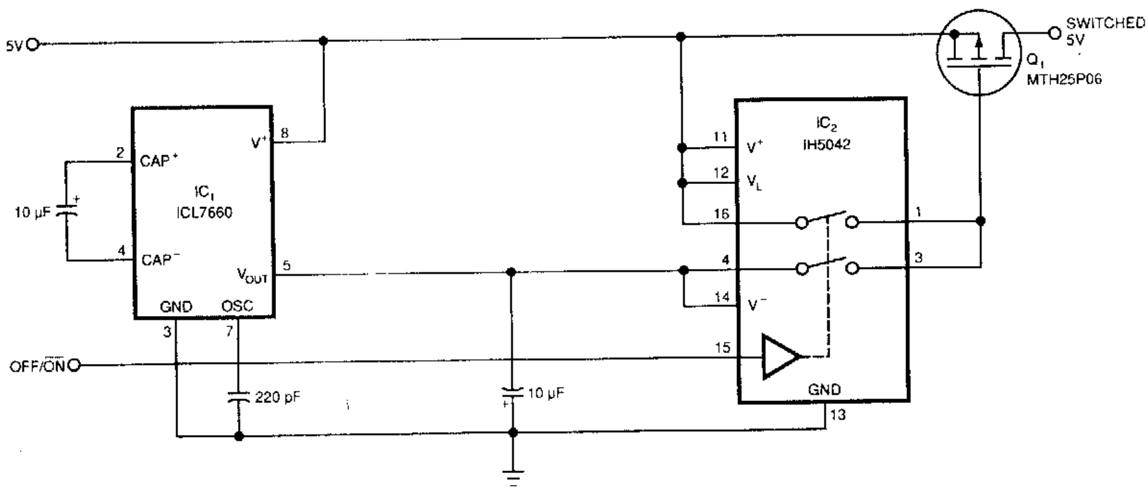
Fig. 61-7

62

Power-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 672. The figure number in the box of each circuit correlates to the entry in the Sources section.

5-V SUPPLY HIGH-SIDE SWITCHER



EDN

Fig. 62-1(a)

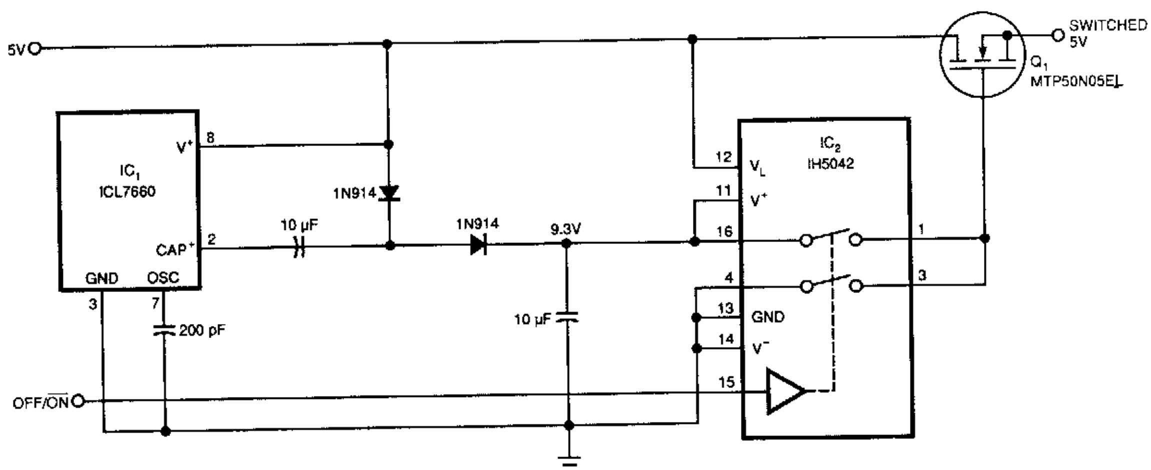


Fig. 62-1(b)

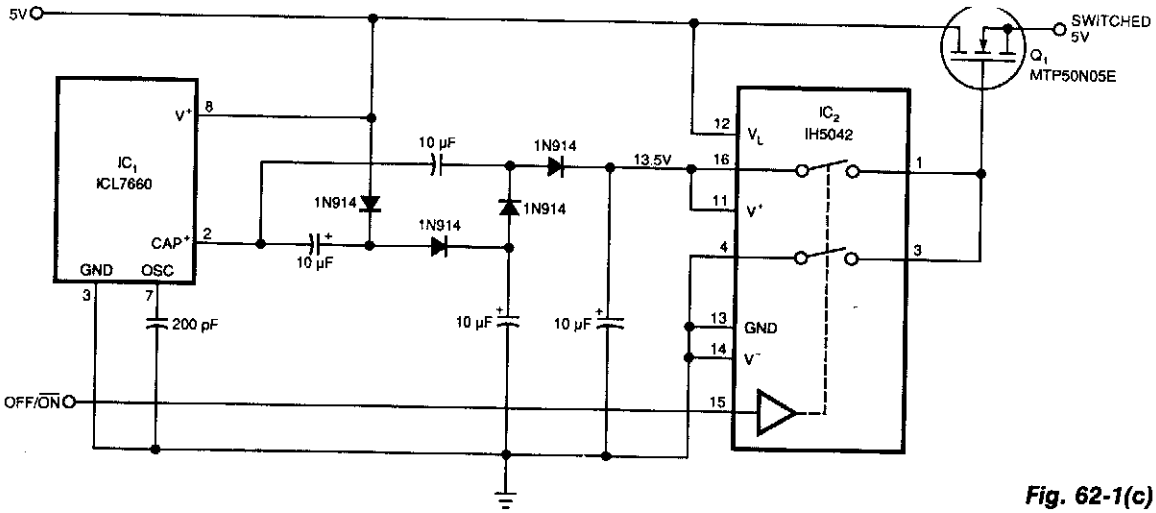


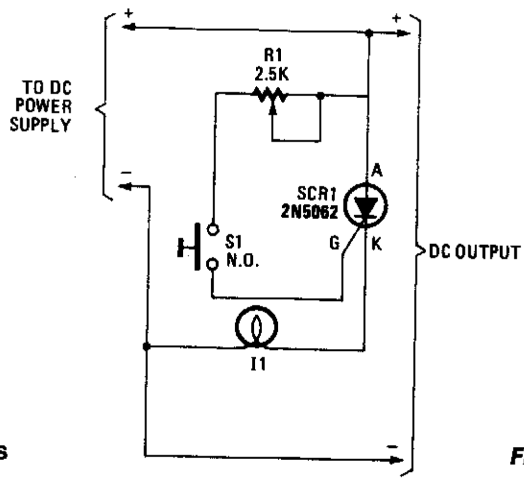
Fig. 62-1(c)

Requiring only 10 μA of quiescent current, the circuit of (Fig. 62-1(a)) produces only 0.1-Ω ON-resistance. IC1 is a charge pump voltage converter to produce a −5-V level, so analog switch IC2 can provide a 10-V swing to MOSFET Q1.

This circuit uses a voltage converter to enable the analog switch to apply a 4.3-V swing to logic level NMOS power transistor Q1. ON resistance is 0.03 Ω typical.

This circuit uses additional stages in the voltage-multiplying circuit to provide a higher gate voltage swing. This would enable the use of a converter for an NMOS switching transistor.

SCR POWER MONITOR

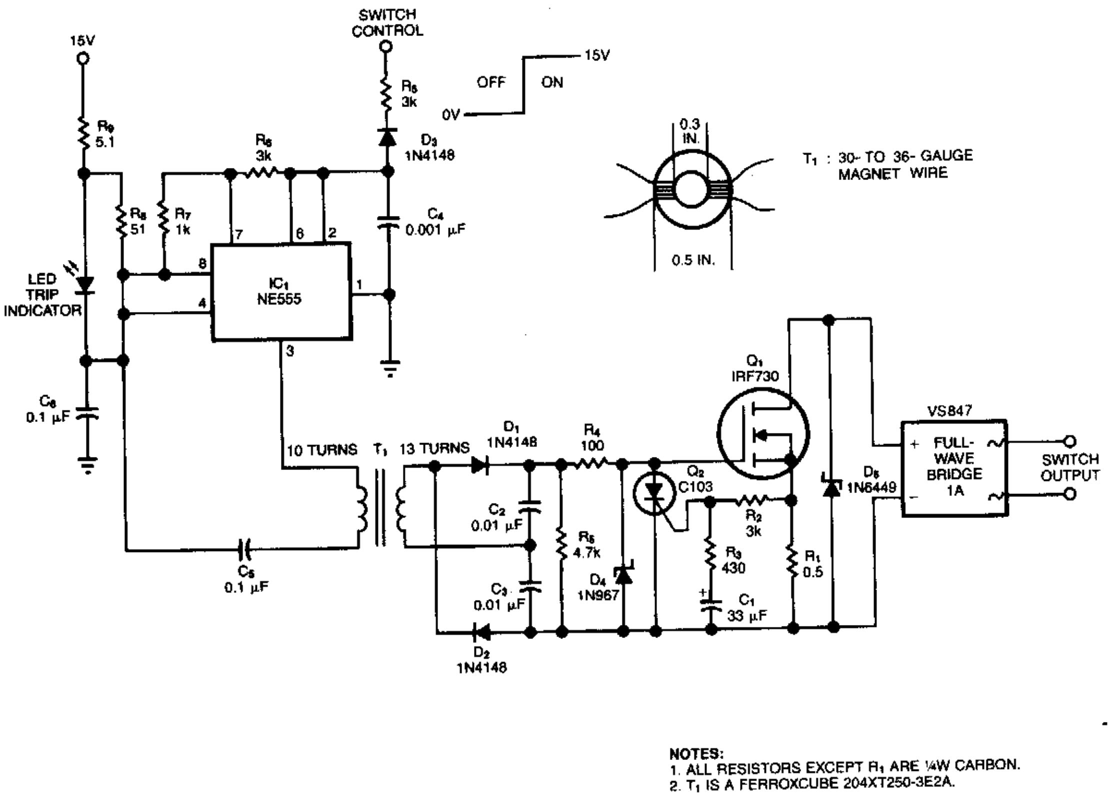


HANDS-ON ELECTRONICS

Fig. 62-2

Pressing R1 lights I1, which will extinguish if there has been a power failure.

POWER MOSFET SWITCH

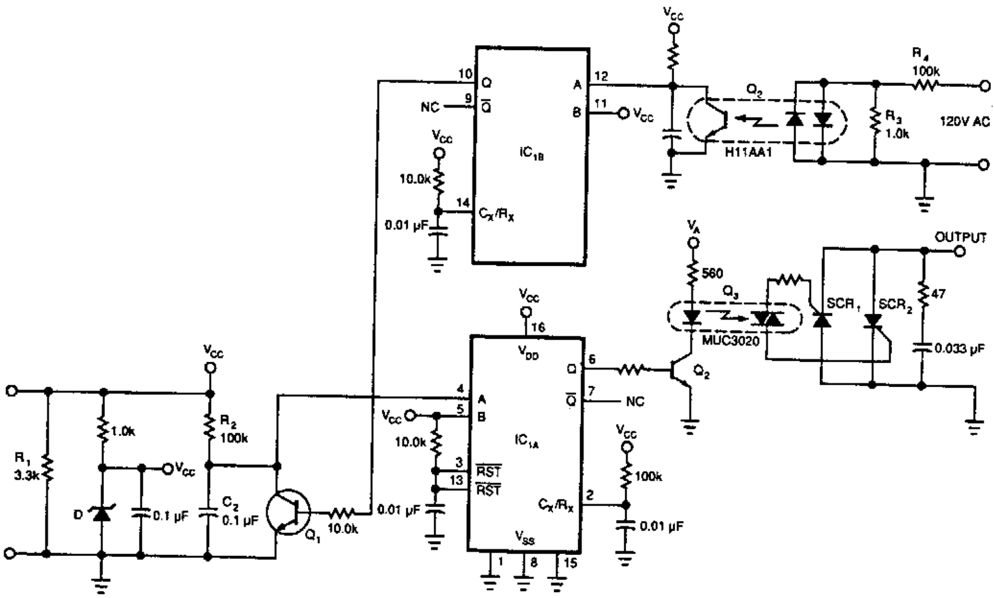


EDN

Fig. 62-3

This solid-state switch senses and interrupts an overcurrent condition within 2 µs. I1 allows the circuit to float. IC1 runs at 150 kHz and full-wave doubler D1/D2 provides 15 V to the gate of Q1. An overcurrent sensed across R1 triggers Q3, removes gate bias from Q1, and opens the circuit formed by the full-wave bridge and Q1. C1 and R3 allow the circuit to handle surges.

CURRENT-LOOP SCR CONTROL

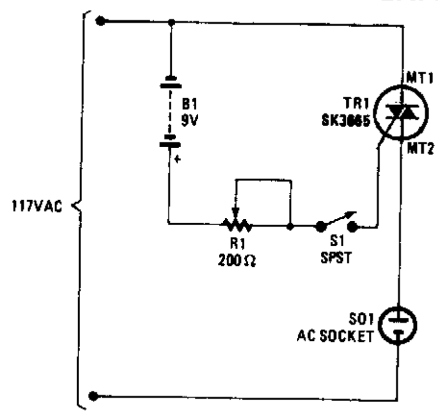


EDN

Fig. 62-4

This circuit allows a 4-to 20-mA current loop to control an isolated SCR drive. IC1A and B are one-shots. Q2 detects zero crossings of the 120 Vac line, which triggers one-shot IC1B. IC1A causes Q1 to discharge C2. When C2 recharges through R2, it triggers IC1A, and the optoisolator and SCR1/SCR2. Triggering of SCR1 and SCR2 is a function of input current, which can control motor speed, light intensity, etc.

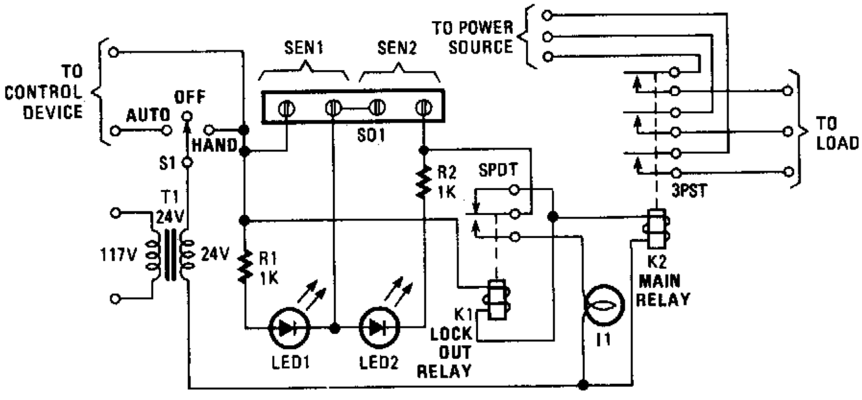
BATTERY-TRIGGERED ac SWITCH



Using this method, a small switch (S1) can control a large ac load. R1 is adjusted for reliable triggering and should be as large as possible.

POPULAR ELECTRONICS *Fig. 62-5*

UNIVERSAL POWER CONTROLLER

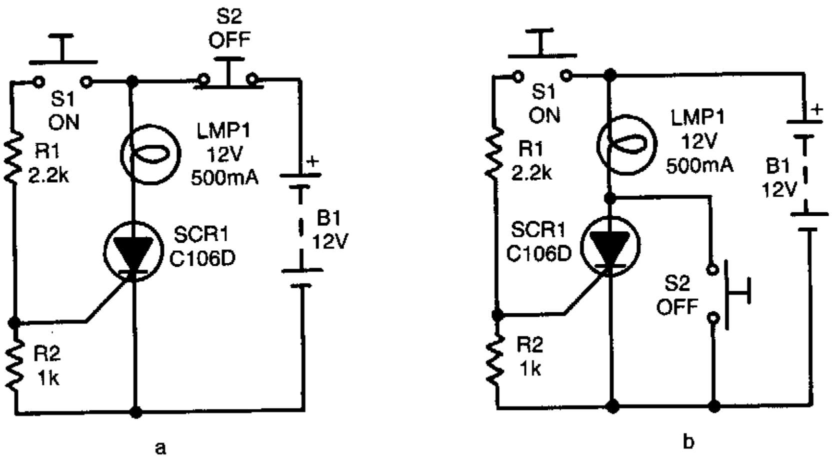


POPULAR ELECTRONICS

Fig. 62-6

Relay K1 has a low-impedance coil and K2 has a high-impedance coil. When a sensor opens, current is routed through the coil of K1. K1 activates, opens its contacts, and prevents a sensor contact reclosure from affecting the circuit. When K1 contacts open, current to the main relay K2 is limited by the impedance of K1. K2 controls power to a load (air conditioner, furnace blower, etc.).

PUSHBUTTON-CONTROLLED POWER SWITCH



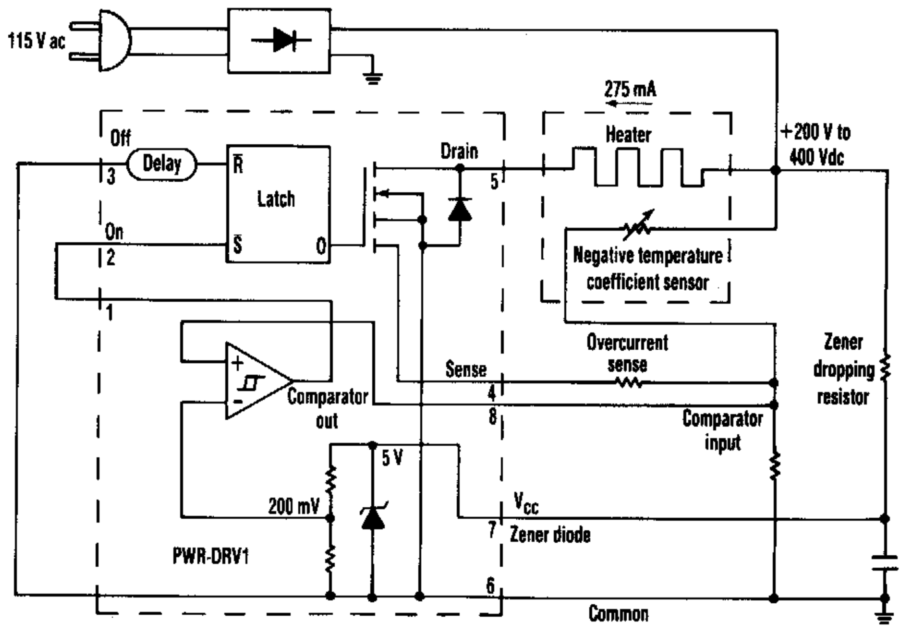
RADIO-ELECTRONICS

Fig. 62-7

In both circuits, the SCR (and thereby the lamp) can be latched on by momentarily closing S1, thereby feeding gate drive to the SC via R1. In both circuits, the gate is tied to the cathode via R2 to improve circuit stability.

Of course, after the SCR turns on, it can be turned off again only by momentarily reducing anode current below the device's I_H value. The SCR is turned off by momentarily opening S2, by using S2 to short the anode and cathode terminals of the SCR momentarily.

BANG-BANG CONTROLLERS

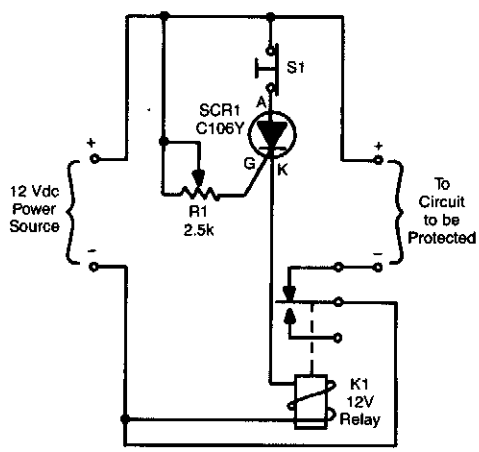


ELECTRONIC DESIGN

Fig. 62-8

Just one chip, the PWR-DRV1 from Power Integrations, builds a "bang-bang" controller that switches 275 mA and runs off the rectified 115-Vac mains. An on-chip zener diode powers the chip from high voltage through a dropping resistor.

SCR OVERVOLTAGE PROTECTOR



HANDS-ON ELECTRONICS

Fig. 62-9

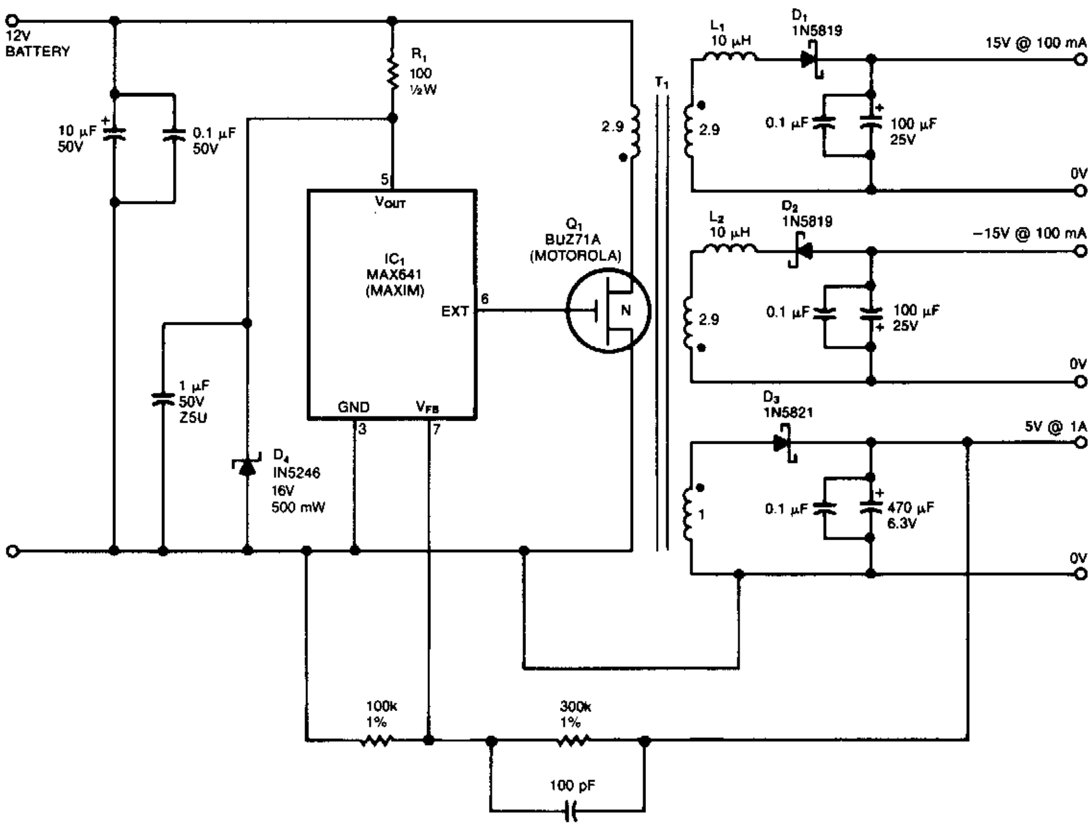
Depending on the setting of R1, when the voltage exceeds a certain amount, SCR1 triggers, which activates K1 and opens the circuit. S1 resets the SCR.

63

Power Supplies (Fixed)

The sources of the following circuits are contained in the Sources section, which begins on page 673. The figure number in the box of each circuit correlates to the entry in the Sources section.

±15-V AND 5-V CAR BATTERY SUPPLY



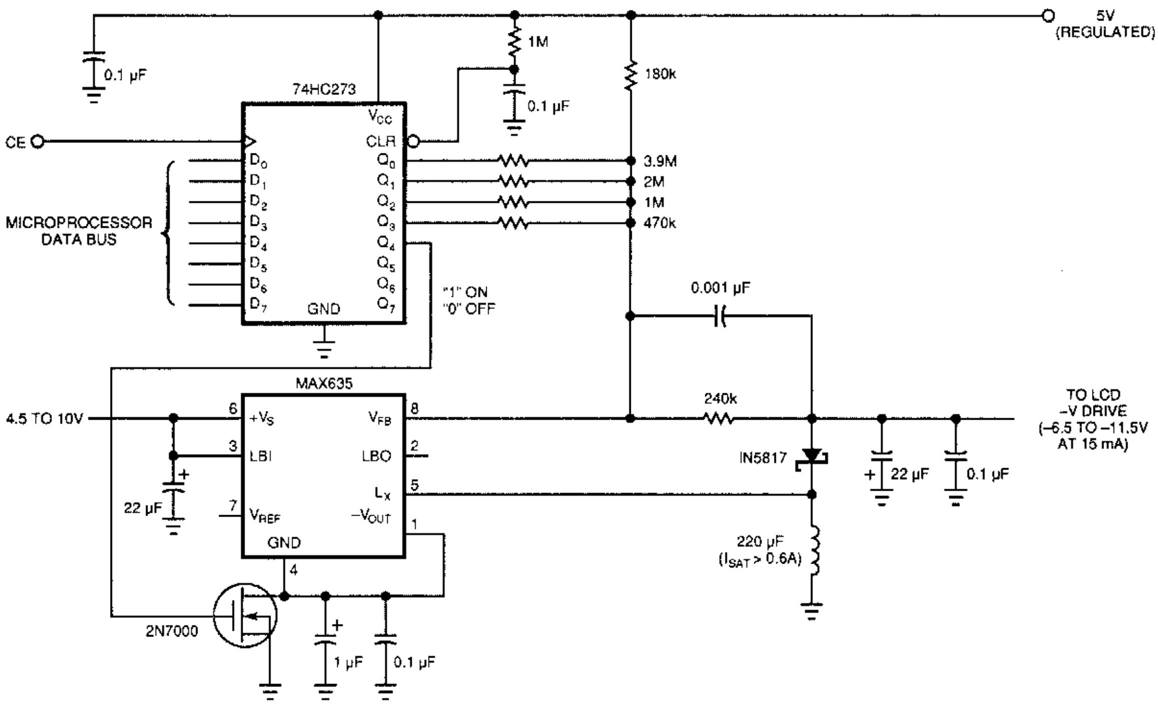
EDN

Fig. 63-1

IC1 is a switching regulator that generates a 45-kHz signal that drives the gate of MOSFET Q1. D1, D2, and D3 are Schottky diodes. The 5-V output is sensed as a reference; feedback to the chip turns off the gate signal to Q1 if the voltage rises above 5 V.

T1 has Trifilar windings that assume about 2% regulation for a 10-to 100-mA load change on the ±15-V supplies. R1/D4 provide overvoltage protection. T1 has a primary inductance of about 21 μH. Core size should allow 4-A peak currents. The turn ratios are 11½ turns each for the 15-V supplies, 11½ turns for the primary, and four turns for the 5-V secondary. The efficiency is about 75%.

SIMPLE LCD DISPLAY POWER SUPPLY



EDN ***Fig. 63-2***

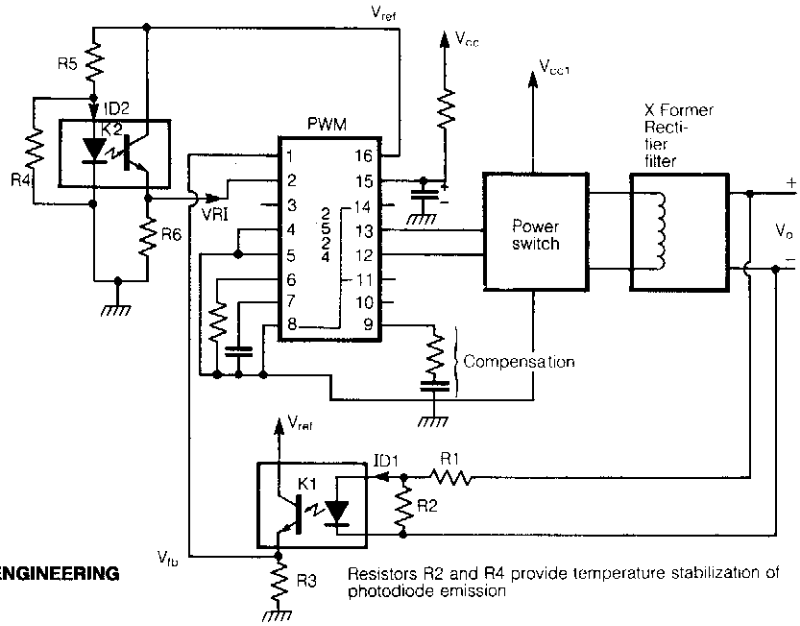
Laptop computers often use large-screen LCDs, which require a variable and a negative supply to ensure maximum contrast. This circuit operates from the system's positive battery supply and generates a digitally variable negative voltage to drive the display.

This figure's switching regulator creates a negative voltage from the battery supply. The microprocessor data bus drives a 4-bit DAC, which in turn varies the actual regulator output from −6.5 to −11.5 V. This arrangement allows a staircase of 16 possible voltages between these limits. The circuit implements the DAC by using the rail-to-rail output-drive capability of a 74 HC-series CMOS gate. A resistor divider network formed by the 240-kΩ resistor, connected to the −V filter capacitor and the resistors, is referenced to the 5-V supply control (the MAX635 regulator).

When the voltage at the V_{FB} pin is greater than ground, the switching regulator turns on. The inductor dumps this energy into the −V filter capacitor. When the voltage at V_{FB} is less than ground, the regulator skips a cycle. The MAX635 regulates the voltage at the junction of the resistor divider to 0 V. Thus, any resistor that the DAC connects to ground (logic 0) will not contribute any current to the ladder. Only the resistors that are at 5 V (logic 1) will be part of the voltage-divider equation.

The entire switching-regulator supply draws less than 150 μA. You can place the circuit in an even lower power mode by interrupting the ground pin. The high-current path is from the battery input through the internal power PMOSFET to the external inductor. Disconnecting the ground connection simply disables the gate drive to the FET and turns off the internal oscillator.

OUTPUT STABILIZER



ELECTRONIC ENGINEERING

Resistors R2 and R4 provide temperature stabilization of photodiode emission

Fig. 63-3

Optically isolated SMPS and dc-to-dc converters face the variance of output voltage owing to the change of transmission characteristics of an optoisolator with (a) temperature and (b) aging. The photo diode emission decays with temperature and time, and causes the output voltage to change. The problem is solved using a homoeopathic principle. An additional optical isolator is used to derive the $+V_e$ input voltage, instead of a conventional potential divider from internally stabilized reference. A scheme is shown using IC2524 as PWM element:

$$1)\ V_{RI}=K_2\times I_{D2}\times R_6 = K_2\left(\frac{V_{\text{Ref}}-V_D}{R_5}-\frac{V_D}{R_4}\right)\times R_6$$

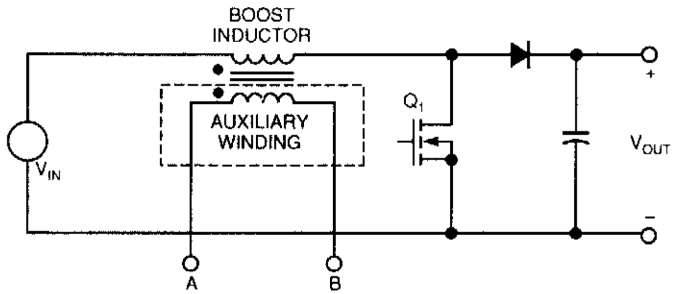
$$2)\ V_{fb}=K_1\times I_{D1}\times R_3 = K_1\left(\frac{V_0-V_D}{R_1}-\frac{V_D}{R_2}\right)\times R_3$$

where V_D = forward drop of a photo diode. In equations 1 and 2, the terms:

$$\frac{V_D}{R_4} \text{ and } \frac{V_D}{R_2}$$

decrease with temperature. Thanks to $-V_e$ temperature coefficient of V_D any changes in K_1 and K_2 as a result of temperature and aging, track each other to maintain the output voltage constant. With proper selection of R_2 and R_4, the output voltages are found to vary by less than 0.01%/°C over a wide temperature range.

AUXILIARY SUPPLY



EDN

Fig. 63-4(a)

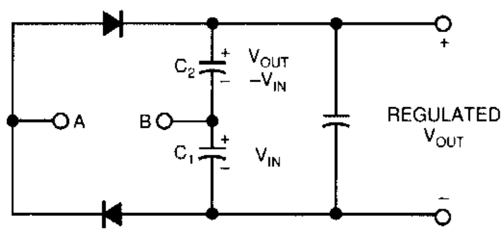


Fig. 63-4(b)

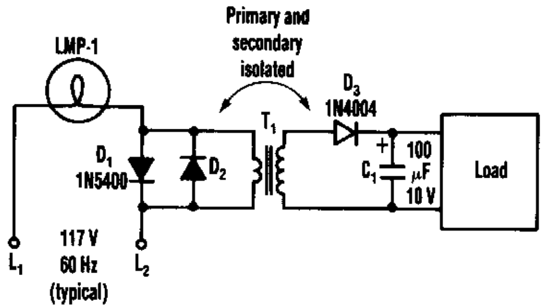
Many power-factor-correction circuits use a boost converter to generate a regulated dc output voltage from the ac line input while forcing the load to draw sinusoidal current, which maximizes the power factor.

This circuit's full-wave rectifier the auxiliary winding's output to completely cancel out line variations and provide a regulated output voltage. The circuit essentially sums the two phases of the boost inductor's voltage to eliminate the 120-Hz components. The regulated output tracks the power-factor-controlled pre-regulator output voltage and it can be used in the corrected output voltage's feedback loop.

An isolated auxiliary winding consists of the desired number of turns wound on the boost inductor. You can vary the exact value of the auxiliary supply's output voltage by adjusting or scaling the auxiliary winding's number of turns. Figure 63-4(b)'s rectifier develops two separate, but individually unregulated voltages, across capacitors C1 and C2. Each of these voltages varies in amplitude at twice the ac-line frequency. When switch Q1 is on, the boost inductor connects directly across the input supply, and a voltage proportional to the instantaneous input voltage develops across capacitor C1.

Once the switch turns off, the inductor voltage reverses and clamps to a voltage equal to $V_{OUT} - V_{IN}$. During this interval, a voltage proportional to $V_{OUT} - V_{IN}$ develops across C2. The sum of these two capacitor voltages produces a regulated auxiliary voltage that is proportional to V_{OUT}. The voltage across the output capacitor equals $V_{IN} + (V_{OUT} - V_{IN})$, which cancels the input-line variations.

ac-TO-dc CONVERTER

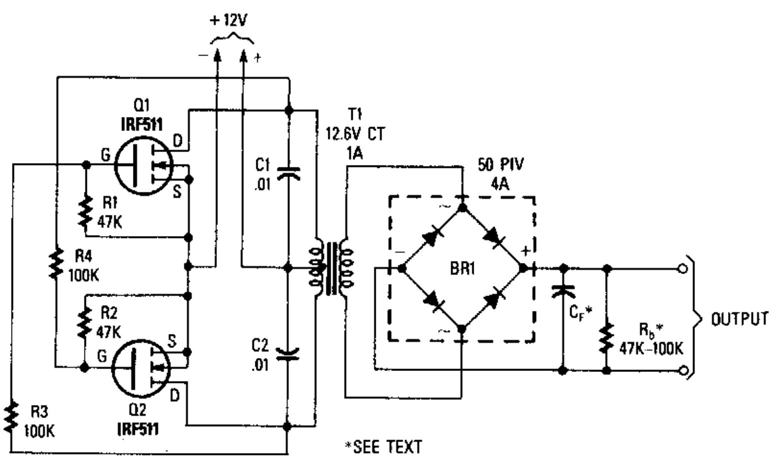


ELECTRONIC DESIGN *Fig. 63-5*

By coupling two back-to-back diodes in series with an ac power circuit, a voltage of about 1.4 Vpp can be obtained. This voltage is useful for exciting the primary coil of a small transformer. The voltage induced in the secondary coil can then be rectified and used to power solid-state control circuits. The forward-voltage drop of the diodes is inherently constant and stable over a wide range of ac-circuit power variations. The resulting voltage developed across the transformer windings is also free from variation that might be caused by changes in the circuit's current or voltage.

In the circuit, a lamp (LMP-1) is connected to the primary ac input line (L1 and L2) through a pair of inverse-parallel-connected power diodes (D1 and D2). As power flows to the lamp, a drop of about 0.7 V is alternatively developed across each of the diodes. This voltage feeds the primary of a small transformer (T1). T1 can be a small 8-Ω to 500-Ω transistor radio output, etc. This will deliver about 11 Vpp across its secondary winding. LMP1 can be a small 120-V lamp of 5 to 25 W, etc.

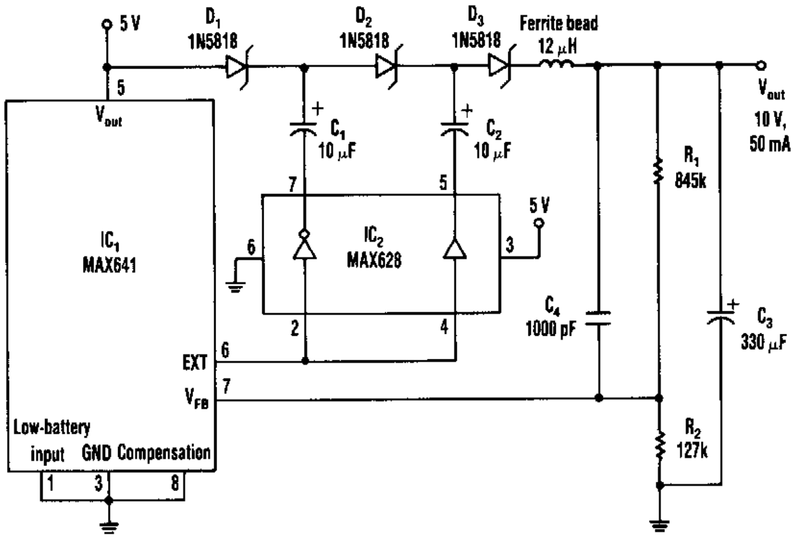
12-V INPUT SIMPLE INVERTER



POPULAR ELECTRONICS *Fig. 63-6*

Using two power MOSFETs, this inverter can deliver ac or dc up to several hundred volts. T1 is a 12.6-V CT to 120-, 240-, or 480-V transformer for 60-Hz application.

REGULATED CHARGE PUMP



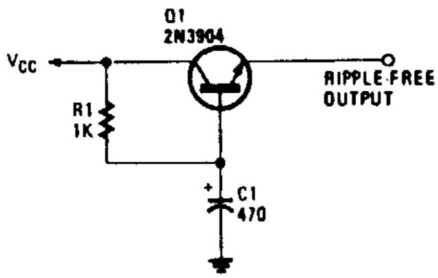
ELECTRONIC DESIGN

Fig. 63-7

The dc-dc converter substitutes a voltage tripler in place of the external inductor and the diode that's typically associated with the switching regulator, IC1. Inverting and noninverting amplifiers in the MOS-FET-driver (IC2) activate a diode-capacitor tripling network (D1 through D3, C1 through C3).

A 50-kHz oscillator residing within IC1 produces the EXT signal (pin 6). IC2 converts this signal into drive signals (180° out of phase) for the tripler. The resulting charge-discharge action in the capacitors recharges C3 toward 10 V every 20 μs. The ferrite bead limits output ripple to about 20-mVpp for a 50-mA load. Conversion efficiency is about 70% for the 5-V input, 10-V output configuration.

SIMPLE RIPPLE SUPPRESSOR

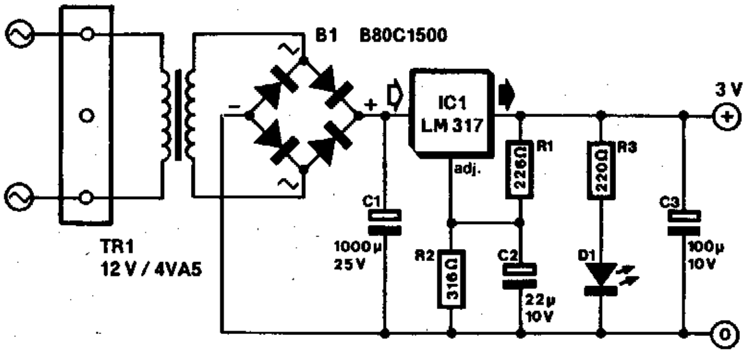


WELS' THINK TANK

Fig. 63-8

This circuit, at times called a *capacitance multiplier*, is useful for suppressing power-supply ripple. C1 provides filtering equal to a capacitor of $(B+1)\ C_1$, where B=dc current gain of Q1 (typically > 50).

3-V POWER SUPPLY FOR PORTABLE RADIOS



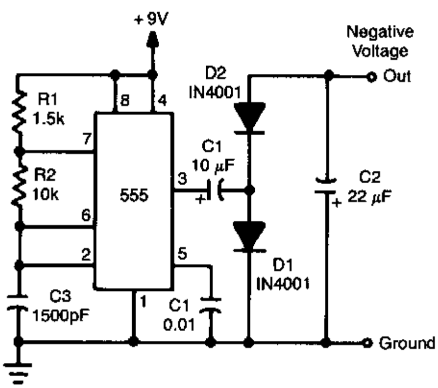
ELEKTOR ELECTRONICS *Fig. 63-9*

Most small portable radios require a 3-V supply, which is normally provided by two AA or AAA batteries. Because rechargeable batteries are an option with many of these radios, most of them are fitted with a charger socket. When such radios are used in a stationary condition (e.g., in the kitchen or in the office), it is useful (and economical) to use the mains-operated supply described here.

The supply is small enough to be fitted inside the radio or in a mains adapter case (less than transformer). Voltage regulator IC1 is adjusted for an output of 3 V by resistors R1 and R2, which are decoupled by C2. Capacitor C3 provides additional filtering. Diode D1 indicates whether the unit has been connected to the mains. The diode also provides the load necessary for the regulator to function properly; in its absence, the secondary voltage of the transformer might become too high when the unit is not loaded.

The transformer should be a short-circuit-proof miniature type, which is rated at 12 V and 4.5 VA. The secondary voltage is slightly higher than needed for a radio, but this reserve is useful when the unit is used with a cassette or CD player. It is advisable to check the output voltage of the unit when it is switched on for the first time before connecting it to a radio or cassette player.

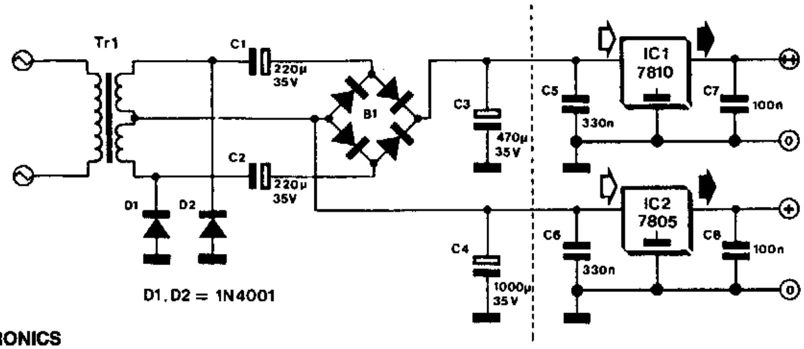
NEGATIVE VOLTAGE FROM A POSITIVE SUPPLY



RADIO-ELECTRONICS *Fig. 63-10*

By using a 555 timer to generate a square wave and voltage-doubling the output, a negative voltage that is almost equal to the positive supply can be obtained. The current available is up to 20 to 30 mA or so, depending on the regulation and voltage needed.

BRIDGE RECTIFIER



ELEKTOR ELECTRONICS

Fig. 63-11

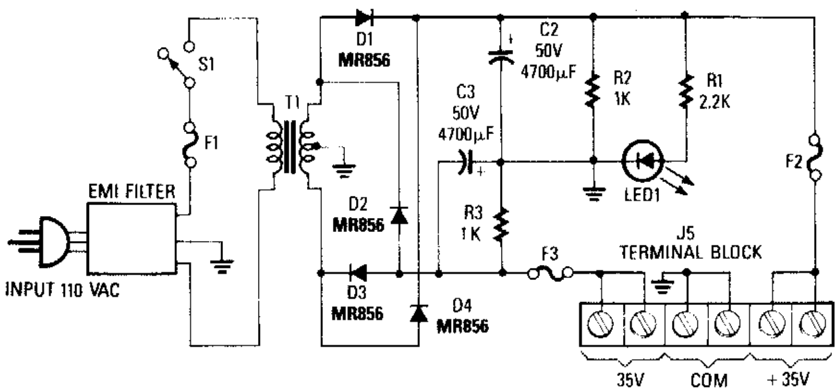
This bridge circuit is intended for those cases where two unequal supply voltages are required. The lower voltage is obtained with the aid of a transformer with symmetric windings and half-wave rectification of the potential across one winding.

For the higher voltage, the potential across both windings is rectified. To that end, the output of the transformer is linked to the bridge rectifier via two electrolytic capacitors that provide isolation of the two direct voltages.

A bonus with this type of circuit is that although the two supplies can be loaded unequally, the currents through the two transformer windings are the same. Thus, the transformer is loaded symmetrically so that its full capacity can be used. Moreover, no unnecessary dissipation is in the voltage regulators.

The load on the lower voltage supply depends primarily on the rating of the transformer. The load on the higher voltage supply is limited by the reactance of C_1 and C_2 ($= 1/2\ \pi\ 50\ C$) and the required minimum output voltage.

±35-V SUPPLY FOR AUDIO AMPLIFIERS

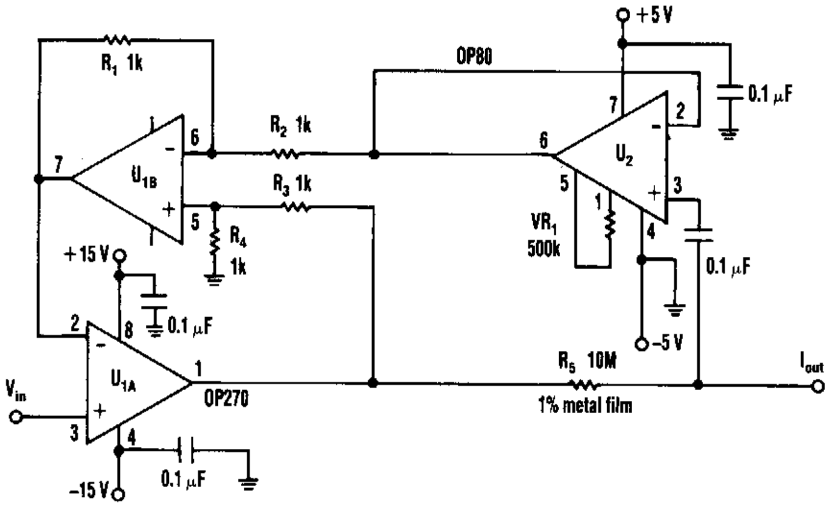


RADIO-ELECTRONICS

Fig. 63-12

This supply will be found useful for operating various transistor AF power amplifiers in the 50-to 100-W output range. T1 = 120 V: 70 V CT at 5 A.

PRECISE LOW-CURRENT SOURCE



ELECTRONIC DESIGN

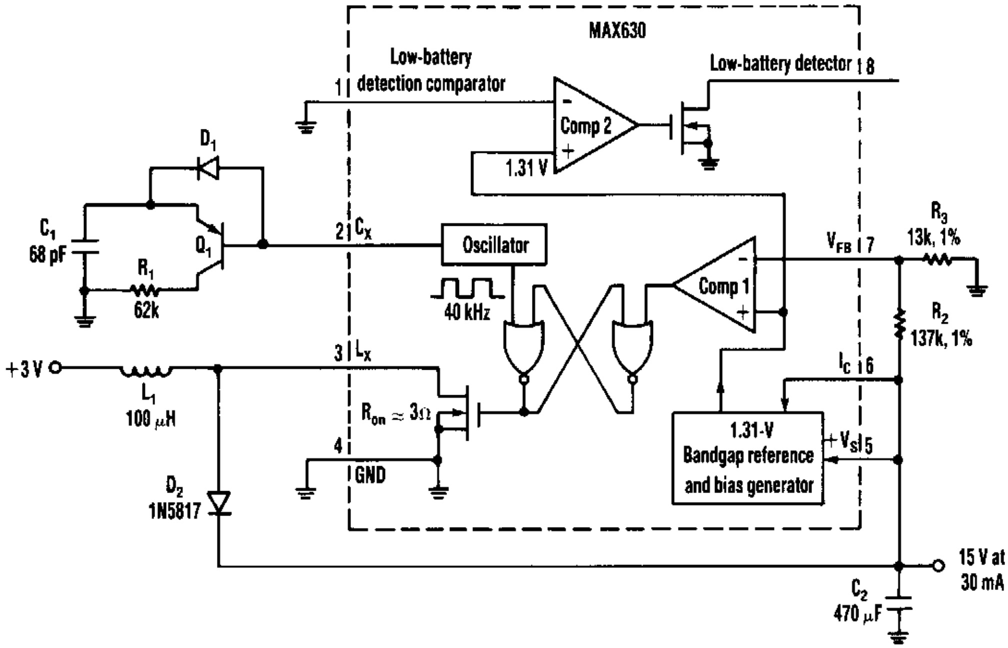
Fig. 63-13

A current source that attains a resolution as low as 10 pA is useful in applications where a precise, low-value current is needed. When the circuit forces current into ground, the output remains within 2% of the ideal current over the ±100-nA range. Over the −4- to +3.5-V compliance range, the error that appears is less than 5%. This accuracy results from using an OP80 op amp from Precision Monolithics in the feedback loop. The OP80 has an I_B of 200 mA typical.

For a given voltage (V_{in}), amp U1A generates an output voltage so that the current through R5 equals V_{in} divided by R_5 (10 MΩ). This current causes a voltage drop across R5, which is sensed by the unity-gain differential amp (U1B and U2). That amp's output is connected to the inverting input of U1A, completing the feedback loop.

The noise in the circuit is of particular concern, especially that produced by resistor R5. The circuit is limited to low-frequency and dc applications as a result of its 400-n V/kHz noise. For applications that don't require the circuit's 10-pA output, lower values of R_5 can be substituted. This will increase the bandwidth.

3- TO 15-V dc-dc CONVERTER

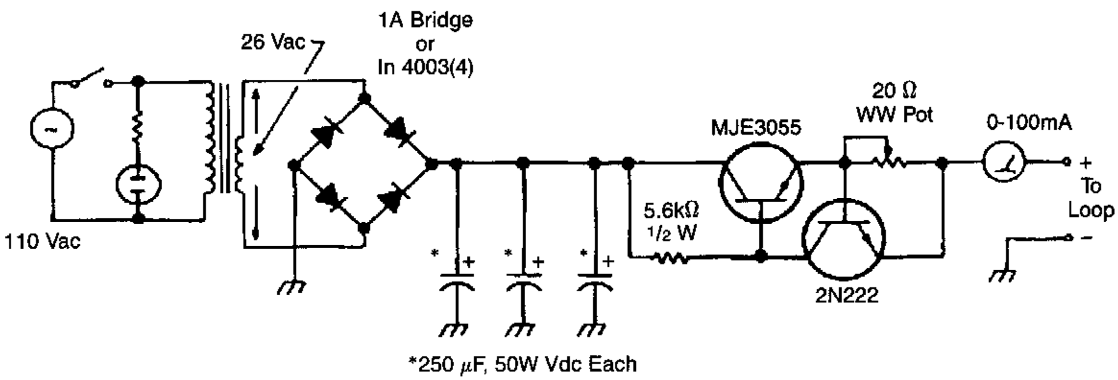


ELECTRONIC DESIGN

Fig. 63-14

This circuit supplies 15 V at 30 mA from a 3-V source. The MAX630 IC is dc-dc converter, Q1 and D1 modify the duty cycle from 50% to 80% to optimize the output power.

CURRENT SUPPLY FOR RTTY MACHINES

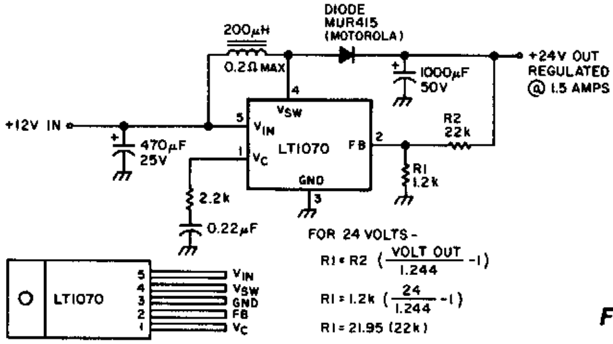


AMATEUR RADIO

Fig. 63-15

Suitable for powering an old Model 15 Teleprinter, this simple power supply uses few parts and is simple to construct. The 20-Ω pot adjusts loop current.

+24-V 1.5-A SUPPLY FROM A +12-V SOURCE

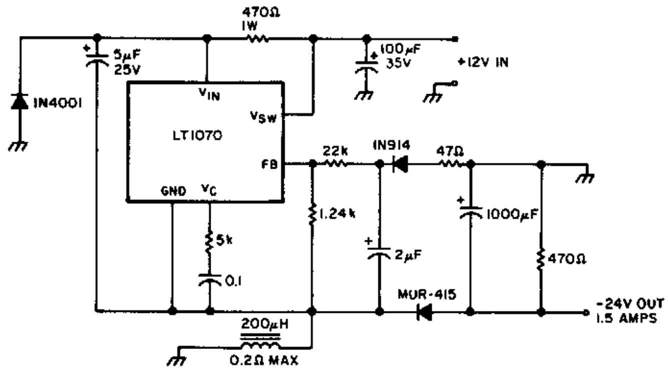


73 AMATEUR RADIO

Fig. 63-16

This switching regulator will produce 1.5 A at 24 V from a 12-V auto battery. It operates as a boost-switching supply using an LT 1070. The diode should be a fast-switching type because this regulator operates above 10 kHz.

NEGATIVE SUPPLY FROM A +12-V SOURCE

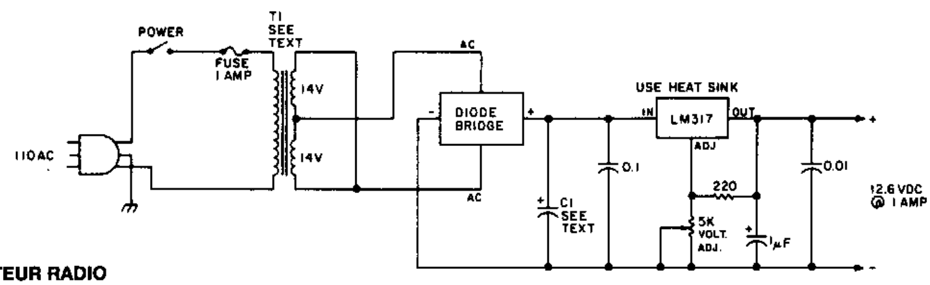


73 AMATEUR RADIO

Fig. 63-17

This switching regulator will produce a −24-V/1.5-A source from a +12-V supply, for applications where a positive-grounded source is necessary.

1-A 12-V REGULATED SUPPLY

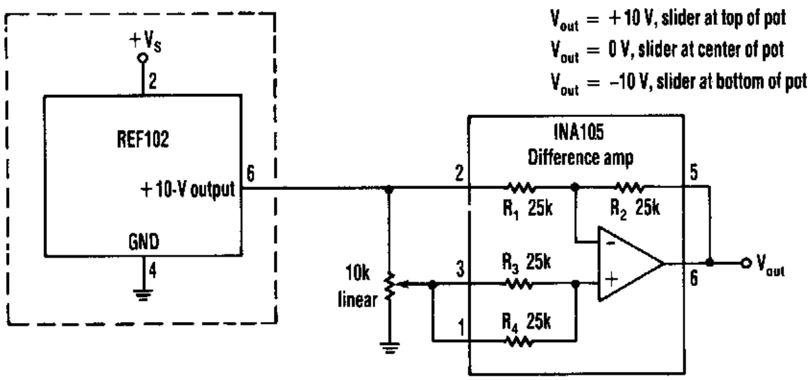


73 AMATEUR RADIO

Fig. 63-18

Using a junked VCR power transformer, this circuit supplies 12 V at 1 A.

POSITIVE AND NEGATIVE VOLTAGE POWER SUPPLY



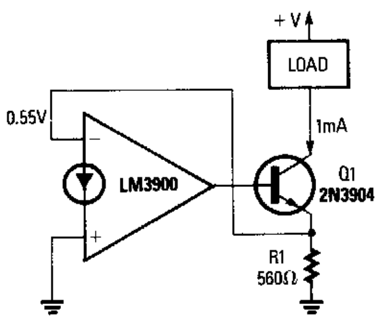
ELECTRONIC DESIGN *Fig. 63-19*

This circuit provides a precision voltage source that can be adjusted through zero to positive and negative voltages, which eliminates reversing connections on the power supply. Also, it is possible to get exactly 0 V, without some offset.

As to how this circuit works, first consider the −1 V/V to +1 V/V linear gain-control amp (see the figure). A Burr-Brown INA105 difference amp is used in a unity-gain inverting amp configuration. A potentiometer is connected between the input and ground.

The pot's slider is connected to the noninverting input of the unity-gain amp; this input is typically connected to ground. With the slider at the bottom of the pot, the circuit is a normal-precision unity-gain inverting amp with a gain of −1.0 V/V ±0.01% maximum. With the slider at the top of the pot, the circuit is a normal-precision voltage follower with a gain of ±1.0 V/V ±0.001% maximum. With the slider in the center, there's equal positive and negative gain for a net gain of 0 V/V. The accuracy between the top and the bottom will usually be limited by the accuracy of the pot.

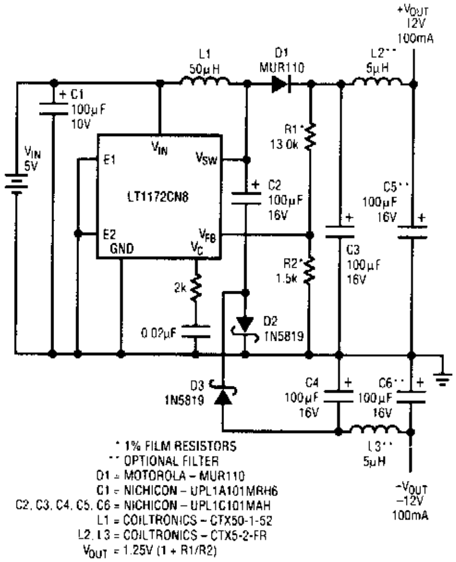
1-mA CURRENT SINK



RADIO-ELECTRONICS *Fig. 63-20*

A fixed current flows through any load that is connected between the positive supply and Q1's collector. The noninverting terminal of the op amp is grounded, and negative feedback flows between the output of the circuit (Q1's emitter) and the inverting terminal. The voltage across R1 is thus equal to the voltage at the inverting terminal (approximately 0.55 V), so a fixed current of about 1 mA flows through the load, Q1's emitter, and R1.

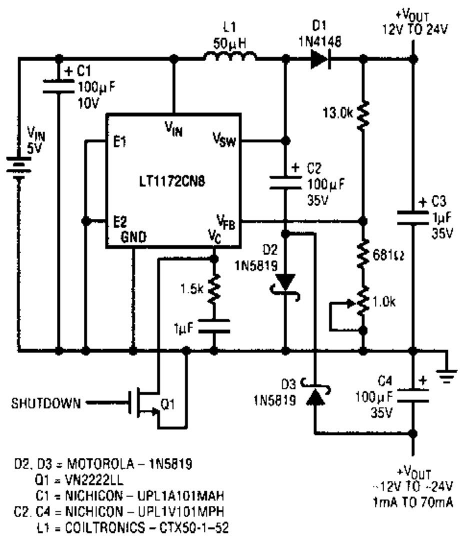
POSITIVE AND NEGATIVE VOLTAGE SWITCHING SUPPLY



LINEAR TECHNOLOGY ***Fig. 63-21***

An LT1172 generates positive and negative voltages from a 5-V input. The LT1172 is configured as a step-up converter. To generate the negative output, a charge pump is used. C2 is charged by the inductor when D2 is forward-biased and discharges into C4 when LT1172's power switch pulls the positive side of C2 to ground.

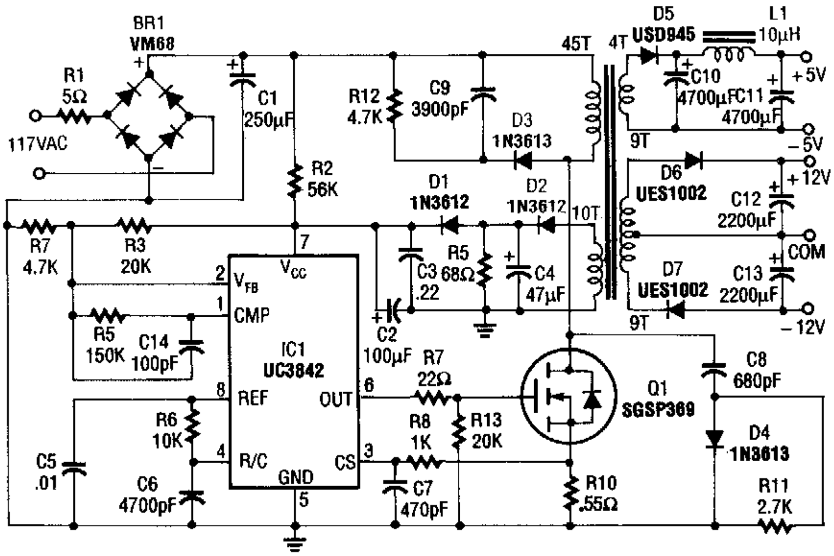
LCD DISPLAY CONTRAST CONTROL POWER SUPPLY



LINEAR TECHNOLOGY ***Fig. 63-22***

The LT1172 is configured as a step-up converter. C2 is charged by L1 and discharges into C4 when the LT1172's power switch goes to ground. Resistor R3 adjusts the output voltage between −12 and −24 V.

5- AND ±12-V ac-POWERED SWITCHING SUPPLY

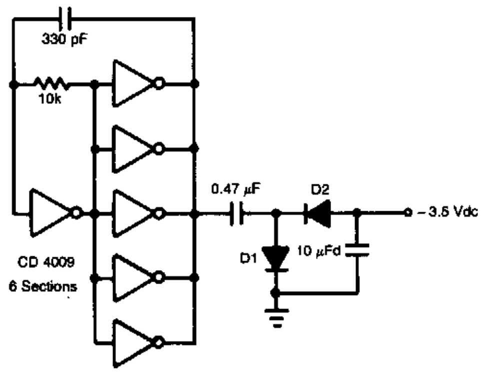


RADIO-ELECTRONICS

Fig. 63-23

This supply uses an SGS-Thomson UC3842 IC in an off-line flyback regulator, providing +5 V at 4 A and ±12 V at 300 mA. This enables a small high-frequency (50 kHz) transformer, to handle large amounts of power that are normally handled by a 60-Hz transformer. Q1 is a 5-A 500-V MOSFET, and the diodes are fast-recovery types. T1 has a 45-turn primary winding of #26 wire. The 12-V windings are each 9 turns of #30 wire, bifilar wound. The 5-V winding is 4 turns of four bifilar #26 wires. The control (feedback) winding is two bifilar, parallel 10-turn, #30 windings. The core is Ferroxcube EC35-3C8 with a 3/8″ center leg.

AUXILIARY NEGATIVE dc SUPPLY FOR BIAS OR REFERENCE APPLICATIONS

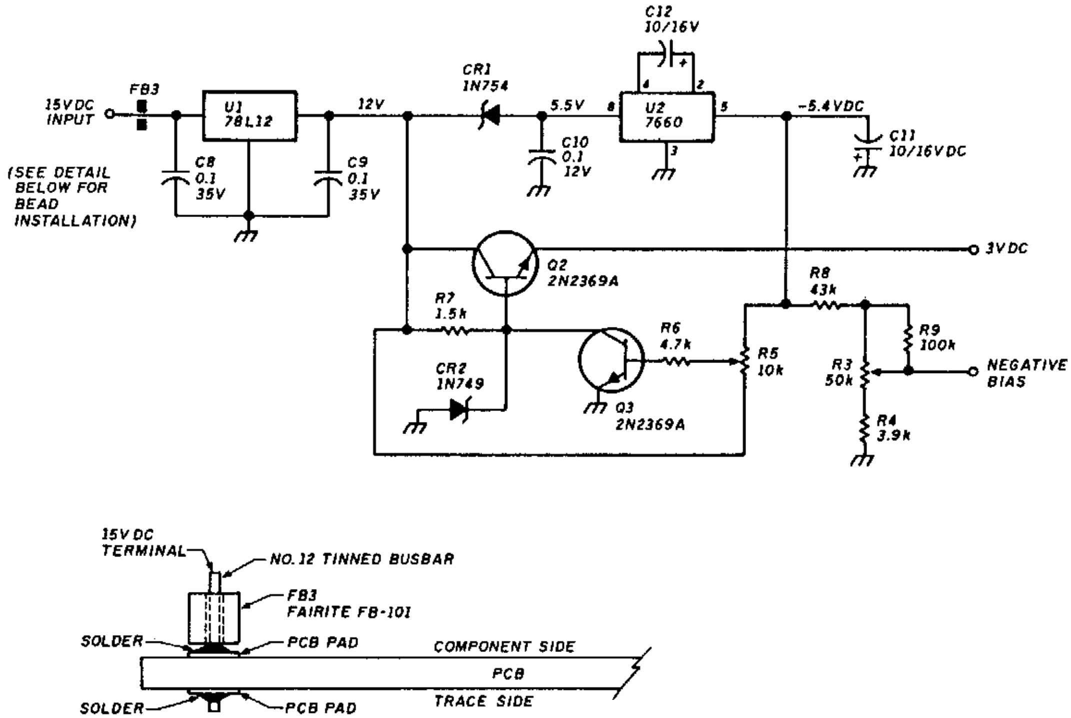


ELEKTOR ELECTRONICS

Fig. 63-24

In this circuit, IC1 (CD4009) is used as a square-wave oscillator at approximately 25 kHz. C1 and R1 set this frequency. C2, D1, D2, and C3 form a p-p rectifier, which outputs about −3.5 Vdc. This circuit should be useful where a small negative dc supply is required, but only positive dc voltages are available.

GASFET POWER SUPPLY

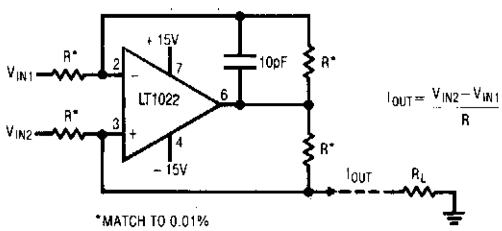


HAM RADIO

Fig. 63-25

Suitable for use with the GASFET doubler in this text, and other similar applications, this supply delivers +3 Vdc and 0 to −3 Vdc bias.

FAST DIFFERENTIAL INPUT CURRENT SOURCE



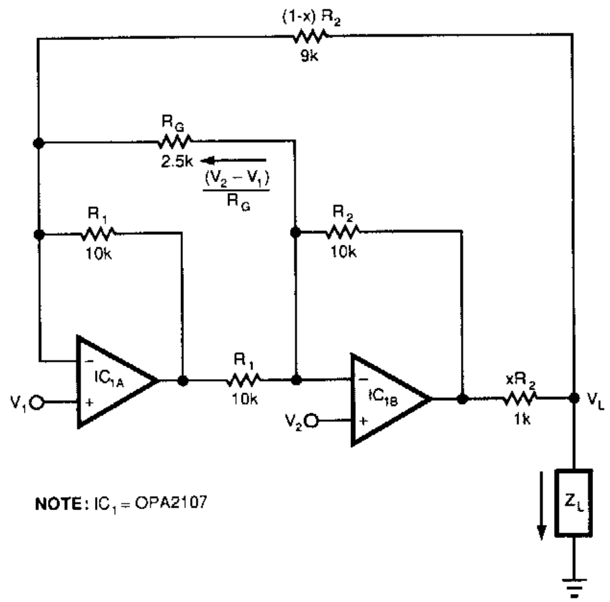
*MATCH TO 0.01%
FULL-SCALE POWER BANDWIDTH
= 1MHz FOR $I_{OUT}R = 8Vp\text{-}p$
= 400kHz FOR $I_{OUT}R = 20Vp\text{-}p$
MAXIMUM $I_{OUT} = 10mAp\text{-}p$
COMMON-MODE VOLTAGE AT LT1022 INPUT = $\frac{I_{OUTp\text{-}p} \times R_L}{2}$

LINEAR TECHNOLOGY

Fig. 63-26

An LT1022 op amp used in this configuration can provide a rapidly switched current source. Use the equations in the figure to select component values.

BOOTSTRAPPED AMP CURRENT SOURCE

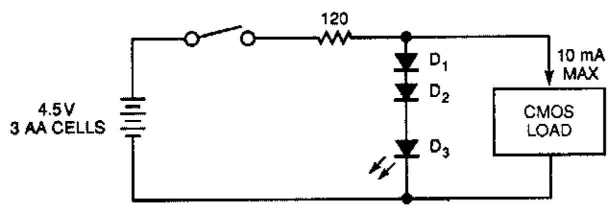


EDN

Fig. 63-27

This circuit responds to the difference between V_1 and V_2. R_G on sets gain. Resistors $XR2$ and (1 – X) R2 produce the bootstrap effect. These two resistors convert the circuit's output voltage to a current. IC1 and IC2 are Burr-Brown OPA2107 or equal.

DIODE CMOS STABILIZER



Voltage regulation

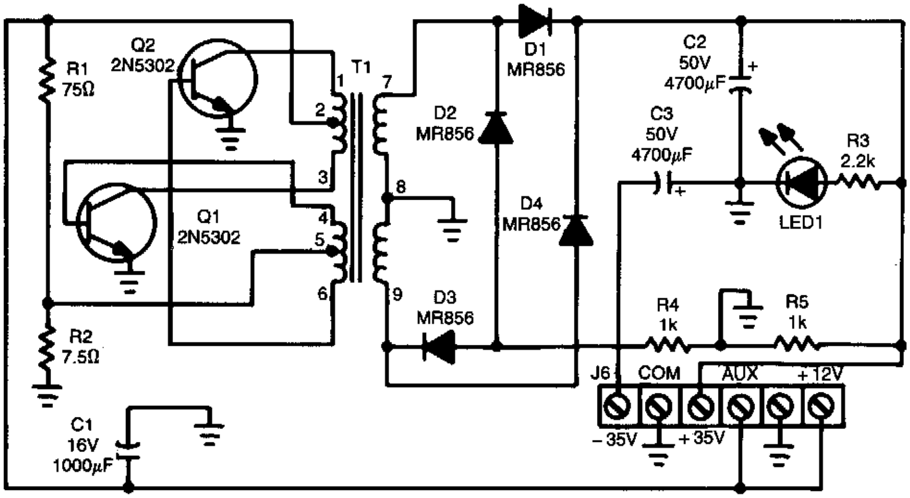
Input	4.5 V	4 V	3.5 V	3.2 V	3 V	2.8 V
Output	3.28 V	3.21 V	3.14 V	3.05 V	2.94 V	2.8 V
LED	on	on	on	on	on	off

EDN

Fig. 63-28

The simple diode network can stabilize the voltage supplied to CMOS circuitry from a battery. D1 and D2 must have a combined forward-voltage drop of about 1.5 V. And D3 is an LED with a forward-voltage drop of about 1.7 V. The table shows the network's output voltage as the battery's voltage declines.

MOBILE ±35-V 5-A AUDIO AMPLIFIER SUPPLY

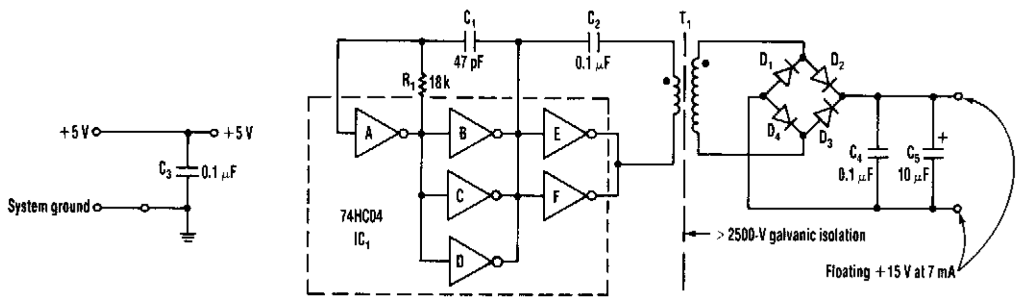


RADIO-ELECTRONICS *Fig. 63-29*

This ±35-V supply uses a two-transistor multivibrator with a toroidal transformer. The transformer core is obtainable from Magnetics, Inc. Specifications for T1 are:

	1 mil tape wound	**1.460″ × 0.915″ × 0.345″**
Core	**Magnetics, Inc.**	**P/N 50029 ID**
Primary	14T CT	#12 AWG
Base Drive	7T CT	#18 AWG
Secondary	19T CT	#12 AWG

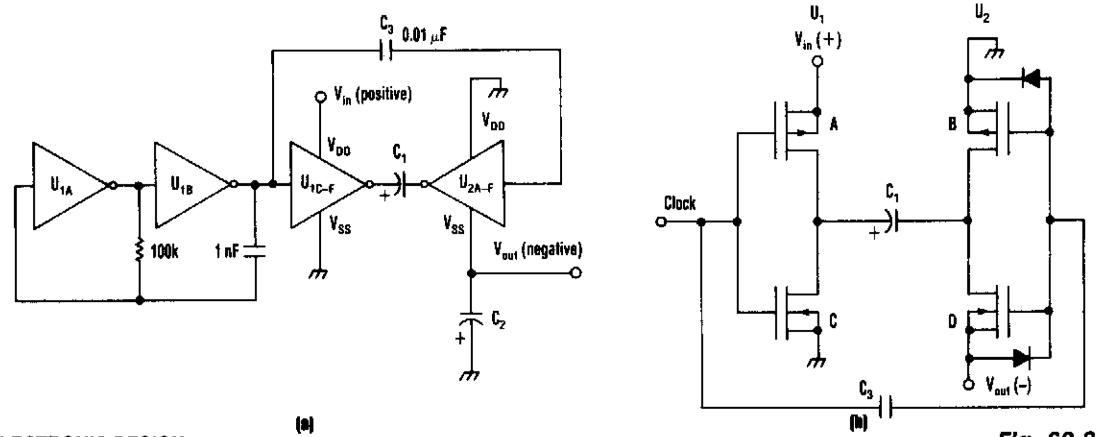
ISOLATED 15-V TO 2 500-V SUPPLY



ELECTRONIC DESIGN *Fig. 63-30*

A dc-dc converter using a 74HC04 drives T1. T1 is a ferrite-core transformer using a Fair-Rite, Inc. P/N 5975000201 ($\mu_o + 5\,000$) and has a 7-turn primary and a 25-turn secondary. Kynar, #30 wirewrap wire is used. With T1, the circuit isolation is good to 2 500 V.

GET NEGATIVE RAIL WITH CMOS GATES



ELECTRONIC DESIGN

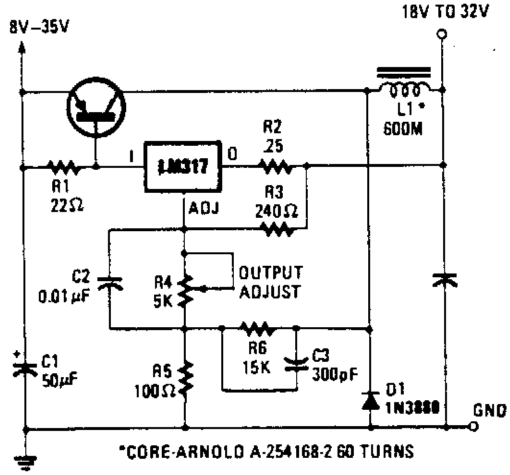
Fig. 63-31

Using a charge pump and oscillator, this circuit uses a 7-kHz oscillator. When the clock is high, C and D are on, grounding the positive side of C1 and making negative voltage available at the V_{SS} terminal of U2. With $C = C_1 + C_2$, the p-p output ripple is:

$$V_{pp}\ \text{ripple} = \frac{V_{in}}{2RFC}$$

Converter output impedance is about 100 Ω and maximum current is 10 mA dc.

3-A SWITCHING REGULATOR



POPULAR ELECTRONICS

Fig. 63-32

This switching regulator uses an LM317 and a pnp switching transistor of 3- to 5-A rating. L1 is wound on a commercially available core.

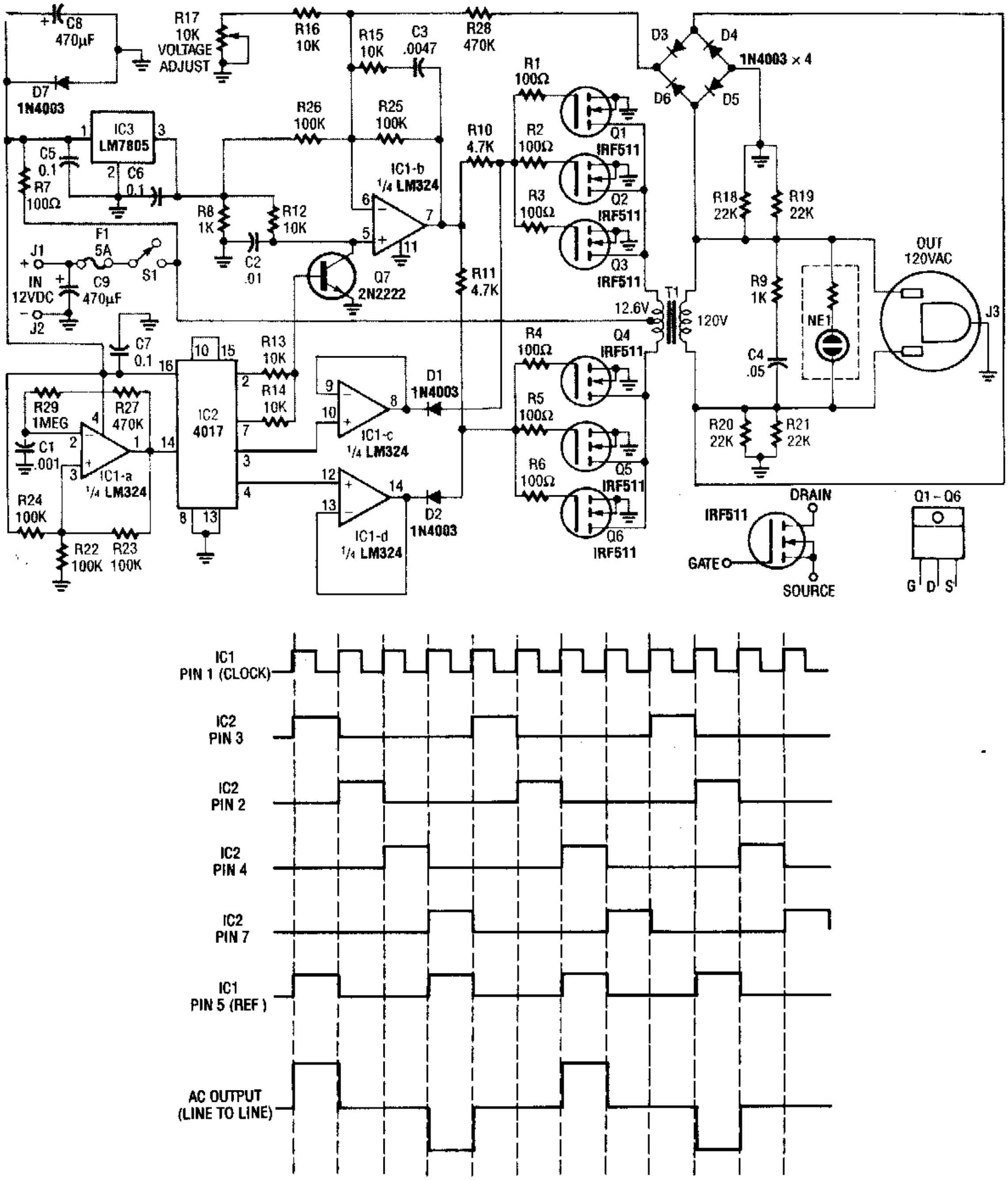
64

Power Supplies (High-Voltage)

The sources of the following circuits are contained in the Sources section, which begins on page 674. The figure number in the box of each circuit correlates to the entry in the Sources section.

40-W 120-Vac Inverter
Cold-Cathode Fluorescent-Lamp Supply
High-Voltage Pulse Supply
High-Voltage Generator
Strobe Power Supply

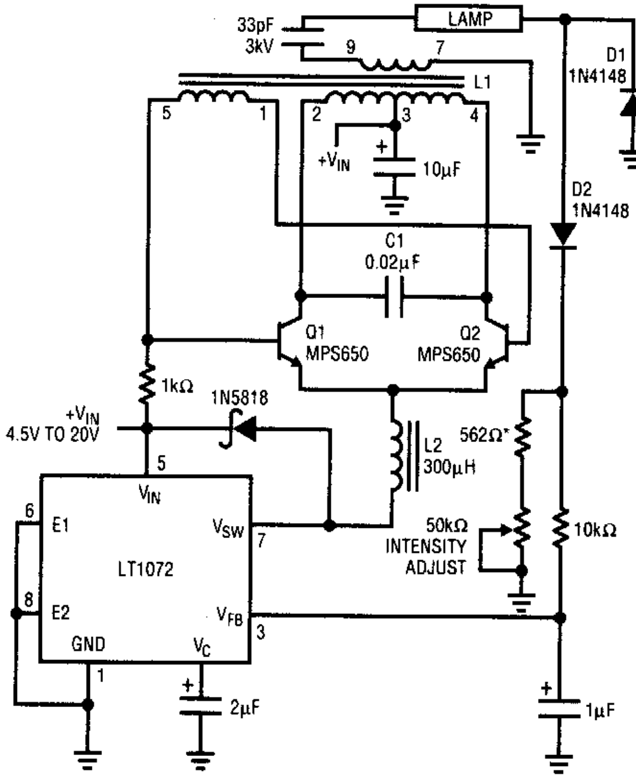
40-W 120-Vac INVERTER



RADIO-ELECTRONICS *Fig. 64-1*

This inverter uses a 12.6-V to 120-V transformer to deliver a quasi-sine wave that has the same rms and peak voltage as a pure sine wave. Q1 to Q6 must be heatsinked. A 1.5″ × 4″ aluminum heatsink was used on the prototype. The transformer should be a 3-A unit. The circuit uses feedback to help regulate the output voltage to 120 Vac. Notice that the output frequency is 75 Hz to avoid saturating the core of T1.

COLD-CATHODE FLUORESCENT-LAMP SUPPLY



C1 = MUST BE A LOW LOSS CAPACITOR.
METALIZED POLYCARB
WIMA FPK 2 (GERMAN) RECOMMENDED.

L1 = SUMIDA 6345-020 OR COILTRONIX CTX110092-1.
PIN NUMBERS SHOWN FOR COILTRONIX UNIT

L2 = COILTRONIX CTX300-4

* = 1% FILM RESISTOR

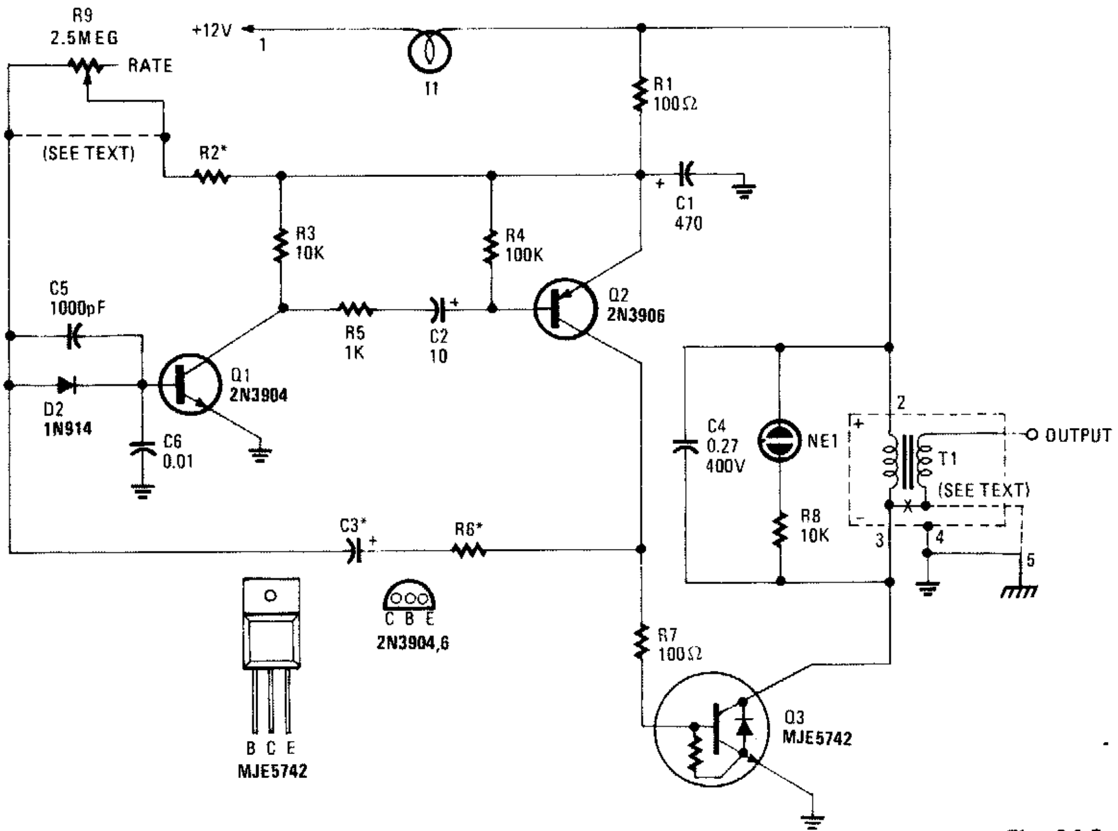
DO NOT SUBSTITUTE COMPONENTS

LINEAR TECHNOLOGY

Fig. 64-2

For back-lit LCD displays, this supply will drive a lamp. LT1072 drives Q1 and Q2, and a sine wave appears across C1. L1 is a transformer that steps up this voltage to about 1 400 V. D1 and D2 detect lamp current and form a feedback loop to the LT1072 to control lamp brightness.

HIGH-VOLTAGE PULSE SUPPLY



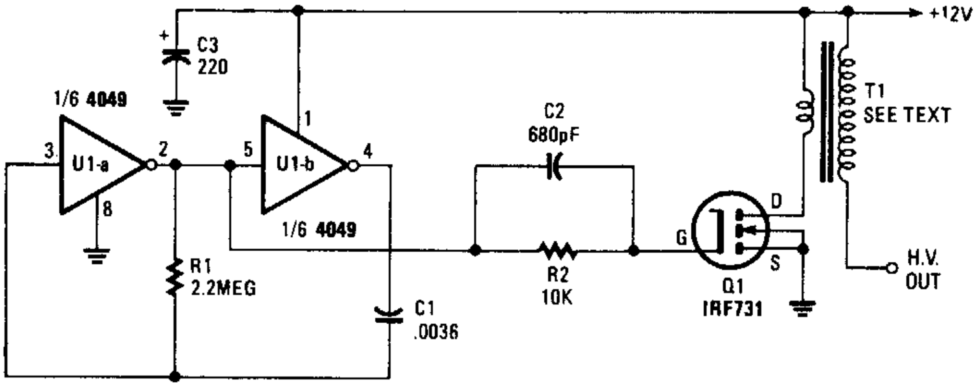
RADIO-ELECTRONICS

Fig. 64-3

This high-voltage pulse supply will generate pulses up to 30 kV. Q1 and Q2 form a multivibrator in conjunction with peripheral components R1 through R6 and C1, C2, C3, C5, C6, and D2. R9 adjusts the pulse repetition rate. R2 should be selected to limit the maximum repetition rate to 20 Hz. I1 is a type 1156 lamp used as a current limiter. R9 can be left out and R2 selected to produce a fixed rate, if desired. Try about 1 MΩ as a start.

Q3 serves as a power amplifier and switch to drive T1 (an automotive ignition coil). NE1 is used as a pulse indicator and indicates circuit operation. Because this circuit can develop up to 30 kV, suitable construction techniques and safety precautions should be observed.

HIGH-VOLTAGE GENERATOR

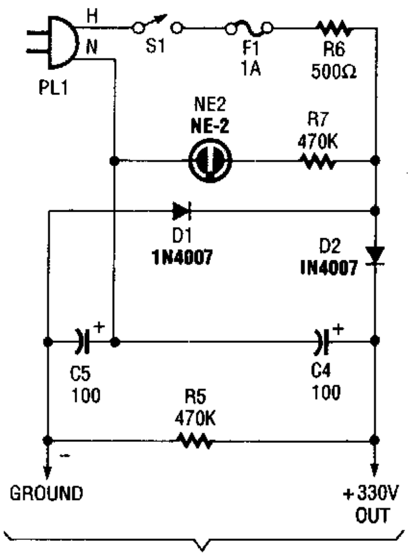


POPULAR ELECTRONICS

Fig. 64-4

A 4049 Hex inverter drives an IRF731 hex FET. The 4049 is configured as an oscillator. Q1 should be heatsinked. T1 is an auto ignition coil.

STROBE POWER SUPPLY



RADIO-ELECTRONICS

Fig. 64-5

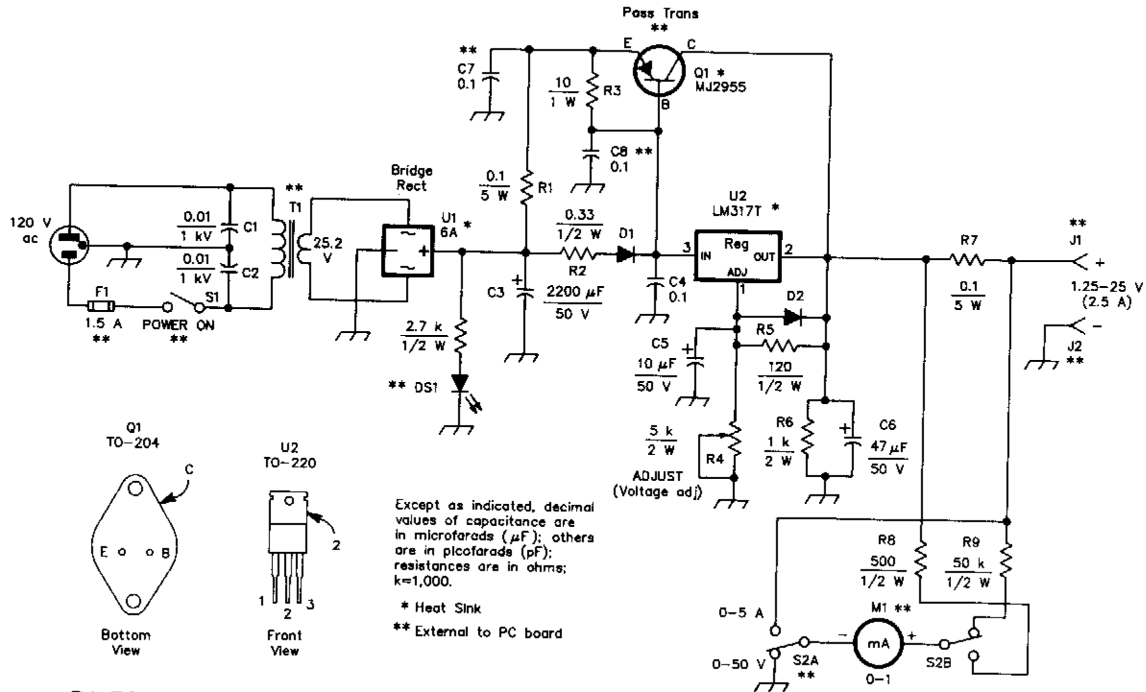
This 330-V power supply is a simple voltage doubler which provides 330 Vdc for a strobe circuit. This supply is not isolated from the power lines and extreme caution is advised.

65

Power Supplies (Variable)

The sources of the following circuits are contained in the Sources section, which begins on page 674. The figure number in the box of each circuit correlates to the entry in the Sources section.

2.5-A/1.25-to 25-V REGULATED POWER SUPPLY



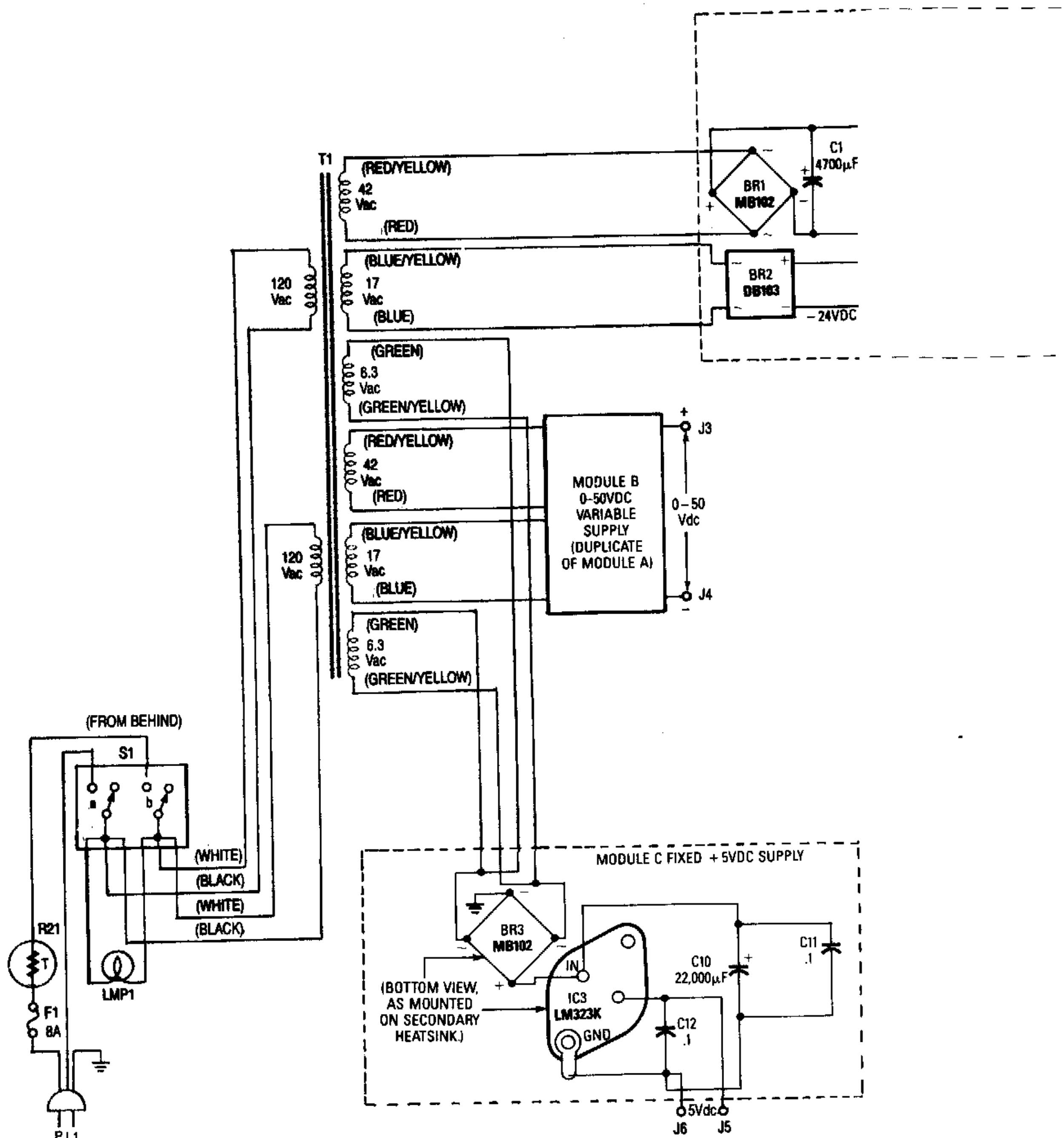
D1, D2—1-A, 100-PIV rectifier diode.
DS1—Red LED.
F1—1.5-A, 3AG fuse in chassis-mount holder.
J1, J2—Standard five-way binding post, one red, one black.
M1—Milliammeter, 0-1 mA dc.
Q1—NPN power transistor MJ2955 (Radio Shack) or equiv device with a +70-V, 10-A, 150-W rating in a TO-204 case.
R1, R2, R7—5-W wire-wound resistor. See Notes 3 and 4 for source. or, use 17 inches of no. 28 enam wire, single-layer wound, on a 10-kΩ, 1-W carbon-composition resistor for R1 and R7. For R2, use 36 inches of no. 30 enam wire on a 10-kΩ, 1-W carbon composition resistor (scramble wound).
R-4—Panel-mount, 5-kΩ, 2-W or 5-W potentiometer, carbon or wire wound (See Note 8).
R8, R9—See text.
S1—SPST toggle switch.
S2—DPDT toggle or rotary wafer switch.
T1—25.2-V, 2.75-A power transformer (see text).
U1—6-A, 200 PIV bridge rectifier with heat sink. See text.
U2—LM317T +1.25- to 30-V, 1.5-A TO-220 regulator. Use an LM317HVK (TO-204 case) for dc output voltage greater than 40. See text.

QST

Fig. 65-1

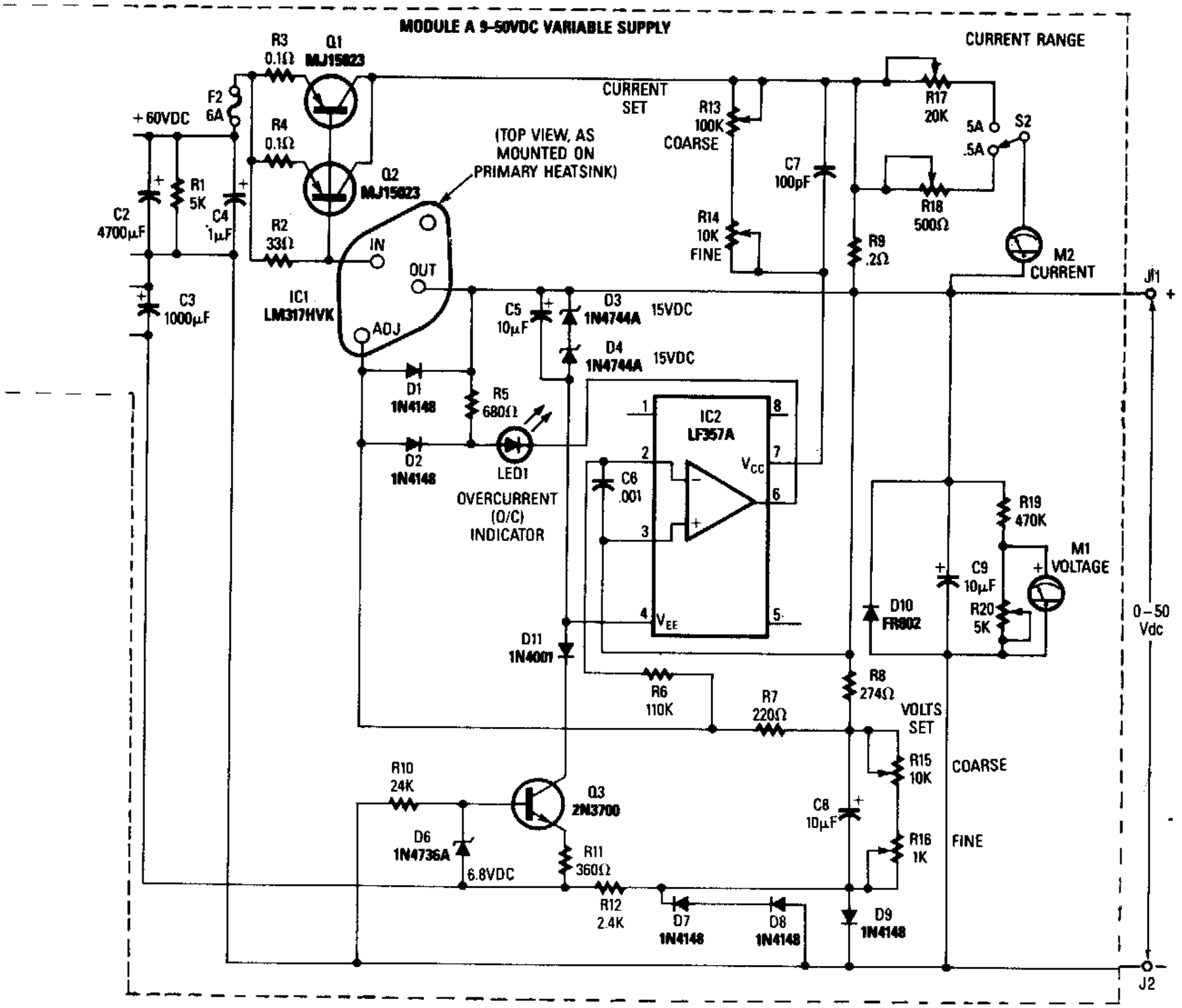
This power supply uses an LM317J adjustable regulator and an MJ2955 pass transistor. Q1 and U2 as well as U1 should be heatsinked. A suitable heatsink would typically be 4″×4″×1″ fins, extruded type, because up to 65 W dissipation can occur. R8 and R9 should be 1% types or selected from 5% film types with an accurate ohmmeter. Capacitors are disc ceramic except for those with polarity marked, which are electrolytic.

DUAL 0-to 50-V/5-A UNIVERSAL POWER SUPPLY



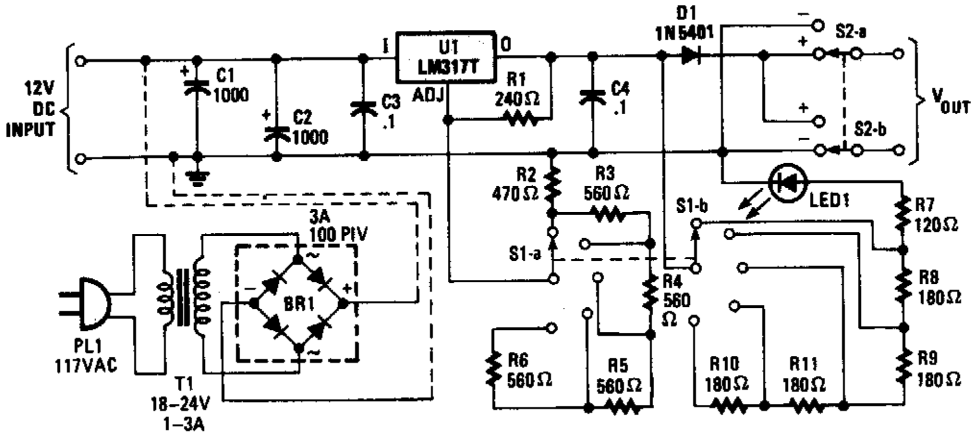
RADIO-ELECTRONICS

Fig. 65-2



T1 has two primaries and six secondaries; the two 120-VAC primaries and 6.3-VAC secondaries are in parallel. Modules A and B are identical; hence, only Module A's parts are called out. Module C is wired point-to-point on the IC3 heatsink.

STEP VARIABLE dc SUPPLY

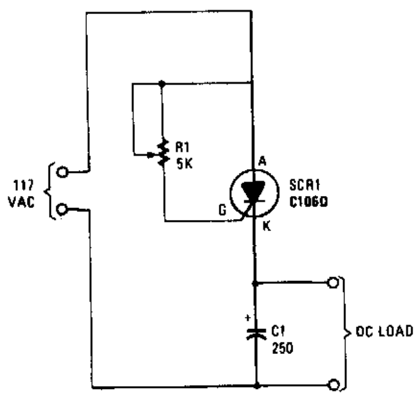


POPULAR ELECTRONICS

Fig. 65-3

Intended as a replacement for generally poorly regulated "wall-type" ac/dc adapters, this circuit offers superior performance to simple, unregulated adapters. Voltages of 3, 6, 9, and 12 V are available. The DPDT switch serves as a polarity-reversal switch. R2 through R6 can be replaced with a 2.5-kΩ pot for a variable voltage of 1 to 12 V. R7 through R10 can be replaced by a fixed resistor of about 1 kΩ if the LED1 brightness variation with output voltage is not a problem.

SCR VARIABLE dc POWER SUPPLY

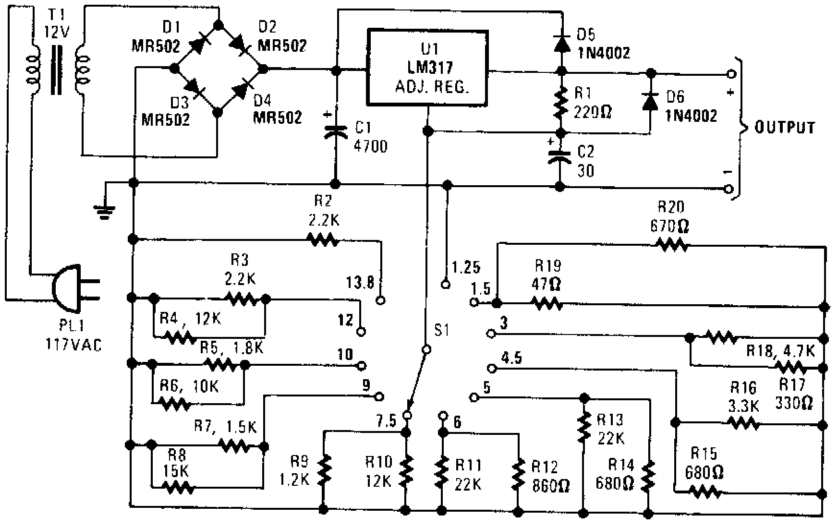


By adjusting the SCR trigger point, a form of phase control can be obtained from an SCR, which produces a dc output, depending on the conduction angle.

HANDS-ON ELECTRONICS

Fig. 65-4

SWITCH-SELECTED FIXED-VOLTAGE POWER SUPPLY



POPULAR ELECTRONICS *Fig. 65-5*

This supply can serve as a battery eliminator for various devices (such as tape recorders, small radios, clocks, etc.).

S1 selects a resistance that is predetermined to provide a preselected output voltage. In this circuit, various commonly used supply voltages produced by batteries were chosen, but any voltages up to the rating of T1 (approximately) can be produced by choosing an appropriate resistor.

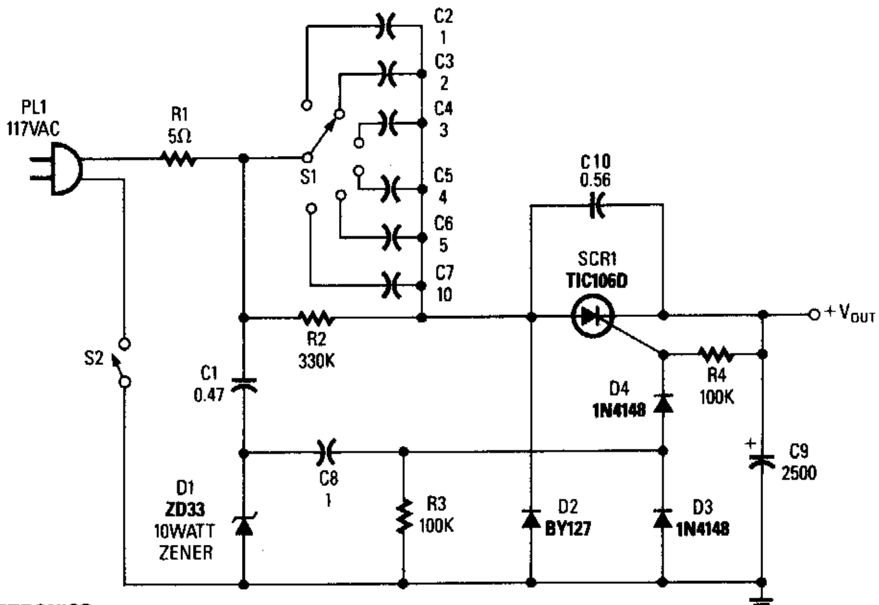
The resistor value is given by:

$$R_X = R_1\left(\frac{V_{OUT}}{1.25} - 1\right)$$

$$R_1 = 220\ \Omega$$

Remember to provide adequate heatsinking for U1.

TRANSFORMERLESS POWER SUPPLY



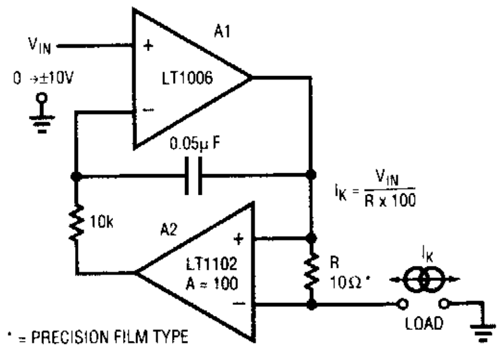
POPULAR ELECTRONICS

Fig. 65-6

By selecting capacitors, various voltages can be obtained from this supply. Notice that C2 through C7 must be nonpolarized capacitors, such as oil-filled or foil types (Mylar) rated for at least 250 Vac.

Warning: This supply is not isolated from the ac mains and presents a serious safety hazard if body contact is made anywhere to this circuit or anything that is powered by it. Use only for applications where contact is avoided or impossible.

VOLTAGE-PROGRAMMABLE CURRENT SOURCE

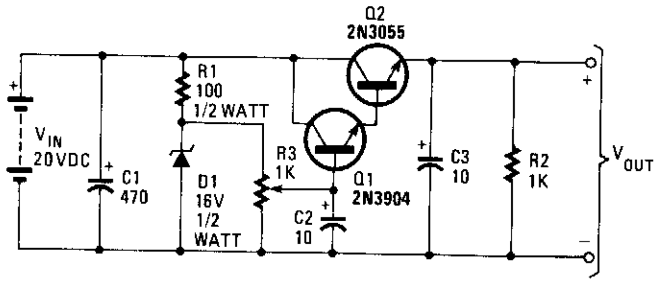


LINEAR TECHNOLOGY

Fig. 65-7

This circuit is a programmable current source in which op amp LT1102 (Linear Technology Corp.) is used in conjunction with LT1006 op amp. A1, biased by V_{in}, drives current through R (10 Ω) and the load. A2 senses this current and controls A1. The 10-kΩ resistor and 0.05-μF capacitor sets the frequency response of the circuit.

SIMPLE DARLINGTON REGULATOR

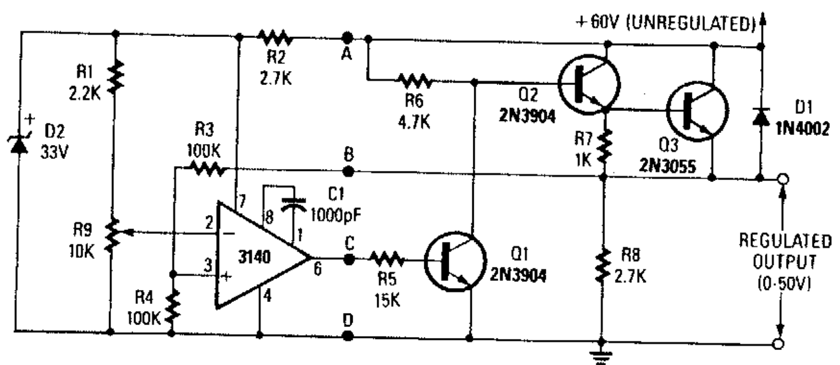


POPULAR ELECTRONICS

Fig. 65-8

A Darlington pair is used as an emitter-follower that produces about 1.2 V less than the wiper voltage of R3. Output voltage for this circuit will range from close to zero to about 14.5 V.

0-TO-50-V VARIABLE REGULATOR



RADIO-ELECTRONICS

Fig. 65-9

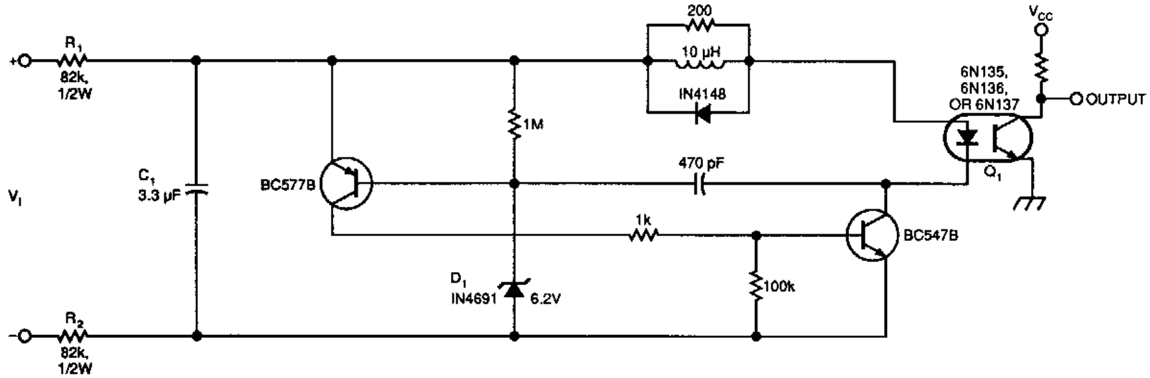
A CA3140 op amp compares the regulator output to a reference voltage, depending on the setting of R9. The output voltage will be nominally twice the voltage between the plus input of the CA3140 and ground. R1 and R9 allow 0 to 50 V.

66

Power-Supply Monitors

The sources of the following circuits are contained in the Sources section, which begins on page 674. The figure number in the box of each circuit correlates to the entry in the Sources section.

ISOLATED VOLTAGE SENSOR



EDN *Fig. 66-1*

A simple voltage-controlled oscillator (VCO), coupled to your instrumentation by an optoisolator, allows you to measure high voltages. The component values suit a 0- to 600-V input range (power dissipation in R1 and R2 set a limit on the input-voltage range). The circuit's linearity is not an issue, because you can linearize its output in software.

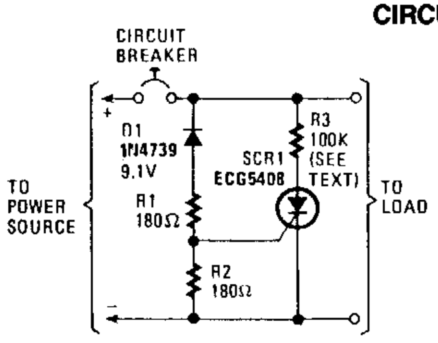
The input voltage (V_1), charges capacitor C1 until zener diode D1 conducts. Then, the zener diode triggers an "avalanche" circuit that discharges C1 into optocoupler Q1. After C1 discharges, the charging cycle repeats. C1 also averages the sensed-voltage level, which thereby provides noise immunity.

The optocoupler's output is a pulse train whose frequency increases with increasing input voltage. To develop a linearizing equation for the circuit, measure its output at two convenient, widely spaced input voltages. Then plug the resulting periods into this second-order polynomial approximation and solve the two simultaneous equations for the two constants, k_1 and k_2:

$$V_1 = [k_1/T20 + (k_2/T) + V_z]$$

V_z is the zener voltage of D1.

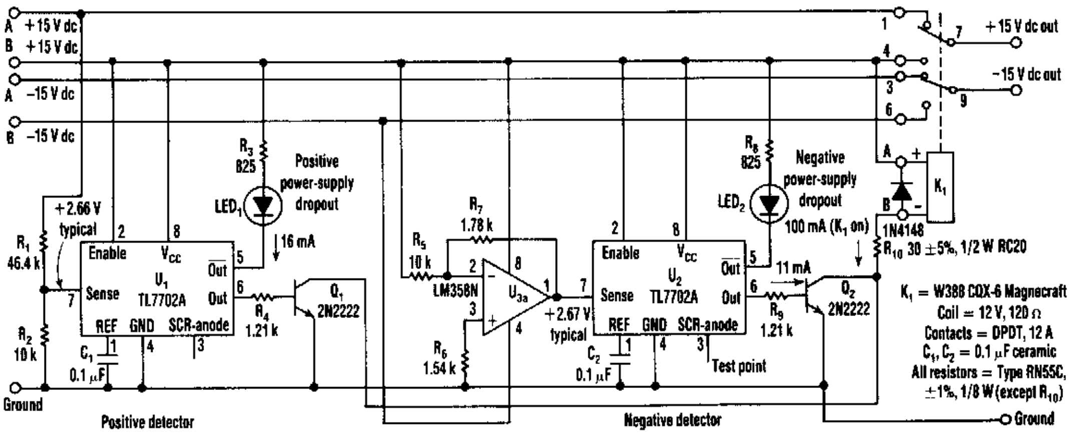
CIRCUIT BREAKER TRIPPER



POPULAR ELECTRONICS *Fig. 66-2*

This tripper is designed to protect against overvoltages. D1 conducts over 9.1 V and triggers SCR1. R3 is chosen to draw enough current to trip the breaker.

BACKUP SUPPLY ACTIVATES BY DROP-IN MAIN SUPPLY

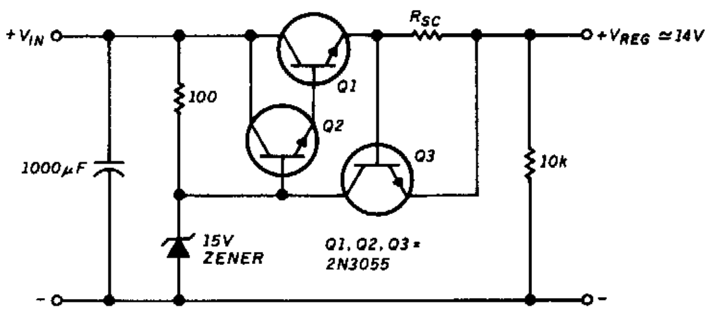


ELECTRONIC DESIGN

Fig. 66-3

A supply monitor using two TL7702A chips monitors the ±15-V supplies and activates the backup supply in case of a voltage drop. Although the chips are intended for use as reset controllers in microprocessor systems, they work well in this application.

CONSTANT-CURRENT TEST LOAD

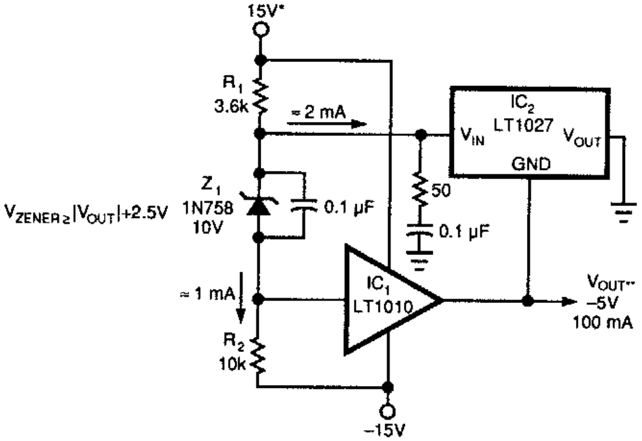


HAM RADIO

Fig. 66-4

This circuit will supply a constant load of 500 mA to 1.5 A. R4 controls the current while R3 provides fine adjustment.

POWER BUFFER BOOSTS REFERENCE CURRENT

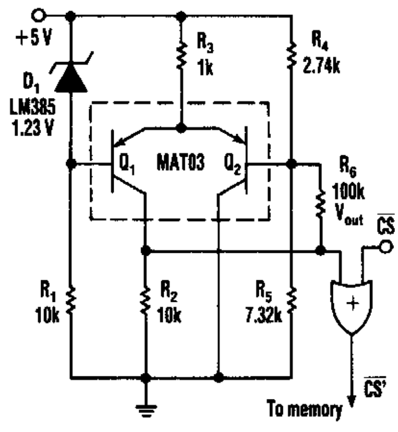


NOTES:
* = BYPASS SUPPLIES WITH 10 µF AND 0.1 µF CAPACITORS
** = OUTPUT WILL OSCILLATE WITH LOW ESR CAPACITORS

EDN *Fig. 66-5*

A method of boosting the output current of a reference and also protecting against overloads is shown in Fig. 66-5. IC1 acts as a power buffer. The LT1027 forces the output of V_{OUT} and ground to be 5 V. The RC damper (50 Ω and 0.1 μF) provides loop stability. The output might oscillate if low ESR capacitors are connected to it, so use aluminum electrolytic or tantalum capacitors instead of ceramic or mylar.

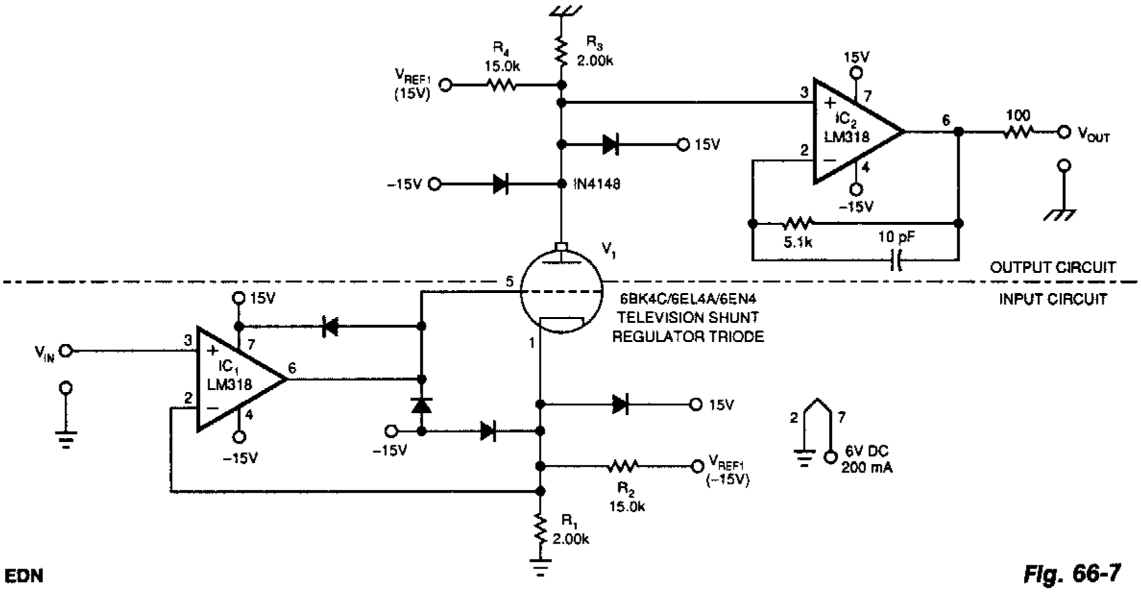
POWER SUPPLY MONITOR/MEMORY PROTECTOR



ELECTRONIC DESIGN *Fig. 66-6*

This circuit detects low-voltage supply conditions, down to 0.6 V. D1 sets the trip point of the circuit. The circuit is useful to protect memory circuits from accidental writes in the event of power-supply low-voltage conditions, which cause other circuits to turn off, etc. Response time is about 700 ns. R6 provides some hysteresis to ensure clean transitions.

TUBE AMPLIFIER ISOLATES HIGH VOLTAGES

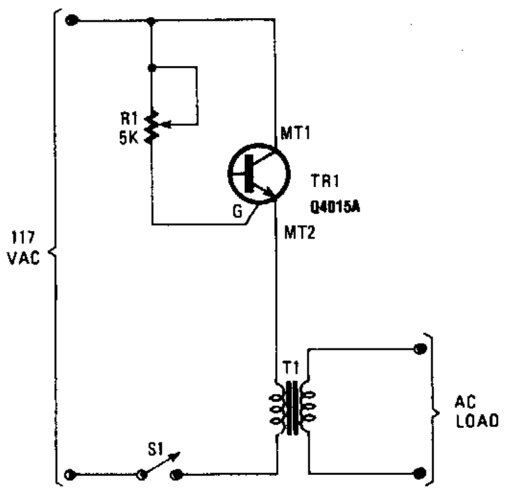


EDN

Fig. 66-7

This amplifier can transfer dc-to 5-MHz signals across a potential difference of 25 000 V. This circuit can be used in CRT displays, high-voltage applications, etc. Notice that the tube must be shielded because the tube will generate X-rays. Typically, about 0.1″-thick sheet metal would be used.

TRIAC ac-VOLTAGE CONTROL

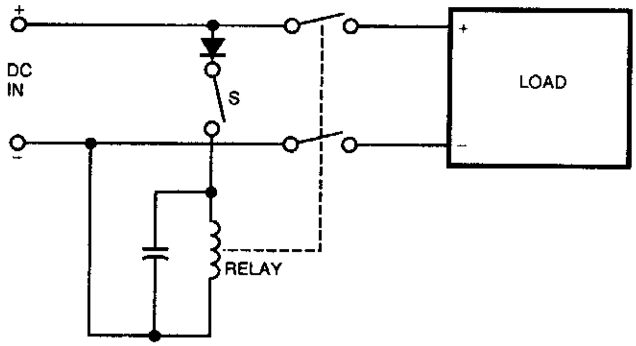


HANDS-ON ELECTRONICS

Fig. 66-8

By using a variable resistor in the gate of TR1, variable conduction angles can be achieved via R1.

POLARITY-PROTECTION RELAY



EDN

Fig. 66-9

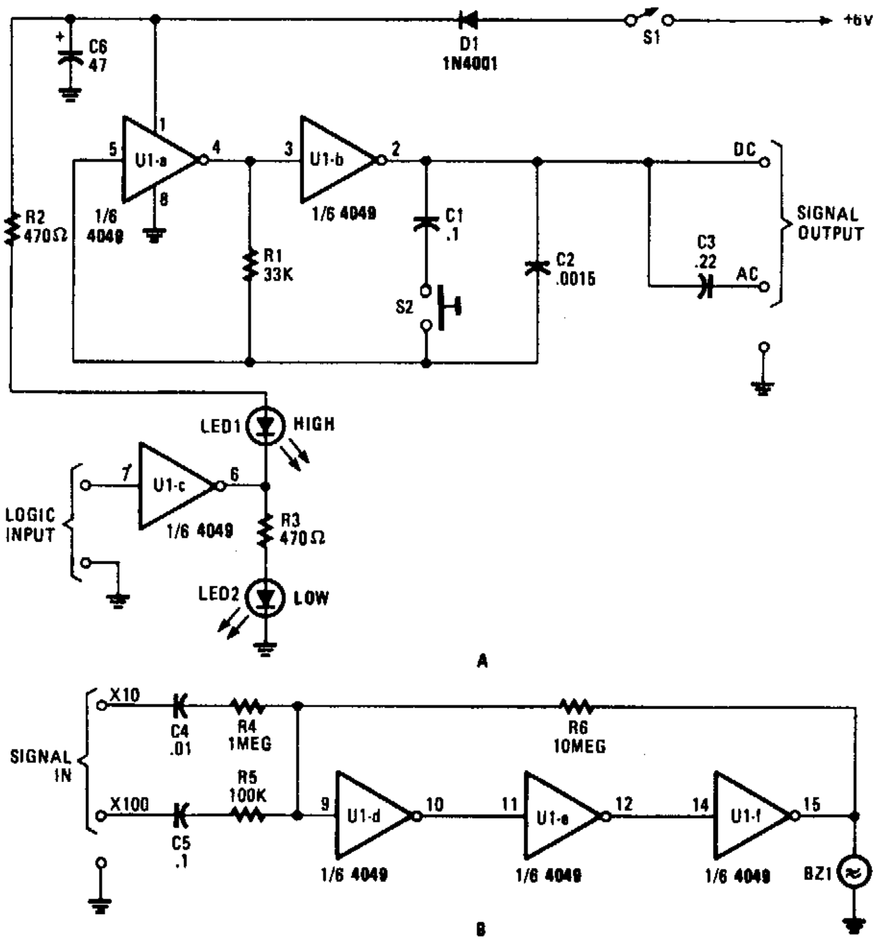
A diode prevents the relay from applying power if polarity is reversed.

67

Probes

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

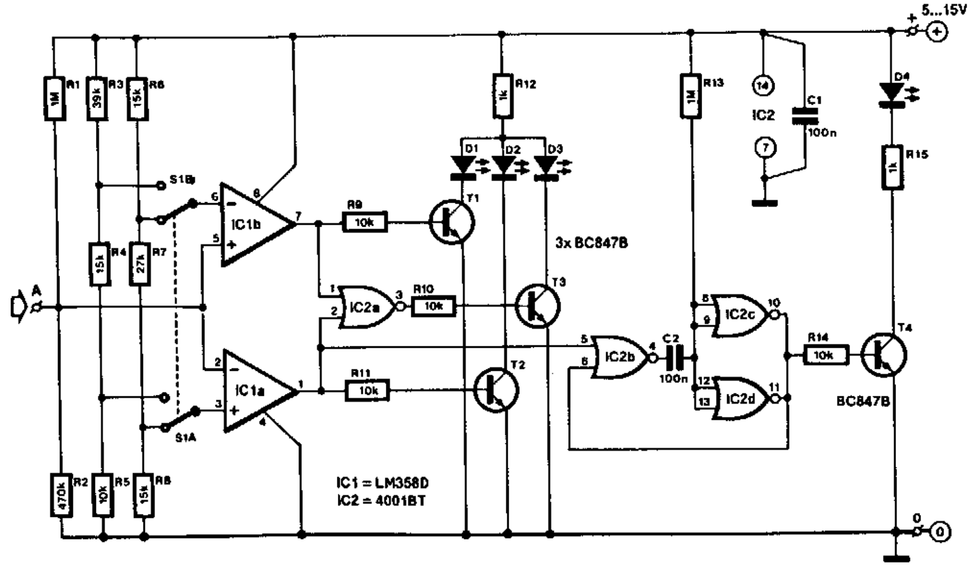
3-IN-1 TEST SET (LOGIC PROBE, SIGNAL TRACER, AND INJECTOR)



POPULAR ELECTRONICS *Fig. 67-1*

This circuit for a test set contains a signal injector (U1A/U1B) and associated components, a logic probe (U1C) and an audio amplifier. S1 selects either 10-kHz or 100-Hz output. U1D, U1E, and U1F form an audio amplifier that drives a piezo sounder element without an internal driver so that it functions as a piezoelectric speaker.

LOGIC TESTER



Parts list

Resistors:
R1, R13 = 1 MΩ
R2 = 470 kΩ
R3 = 39 kΩ
R4, R6, R8 = 15 kΩ
R5, R9, R10, R11, R14 = 10 kΩ
R7 = 27 kΩ
R12, R15 = 1 kΩ

Capacitors:
C1, C2 = 100 nF

Semiconductors:
D1, D2 = LED, 3 mm, green
D3 = LED, 3 mm, red
D4 = LED, 3 mm, yellow
T1, T2, T3, T4 = BC847
IC1 = LM358
IC2 = 4001

Miscellaneous:
S1 = sub-miniature switch, 2 make-before-break contacts

ELEKTOR ELECTRONICS

Fig. 67-2

The input consists of two comparators that operate with different reference voltages supplied by separate potential dividers. Divider R3/R4/R5 provides a voltage of about 40% of the supply voltage, U_{CC}, to pin 6 of IC1B and one of about 16% of U_{CC} to pin 3 of UC1A. When $U_{CC}=5$ V, these voltages are exactly the thresholds (0.8 and 2.0 V) of the TTL comparators.

Similarly, divider R6/R7/R8 provides voltages of 23% of U_{CC} and 73% of U_{CC} to pin 3 of IC1A and pin 6 of IC1B respectively; these levels correspond to the standard threshold for CMOS comparators.

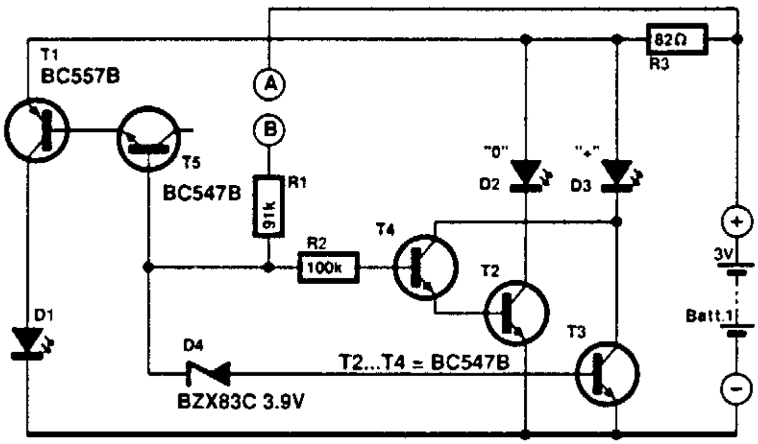
The voltage to be measured U_a, is applied to pin 5 of IC1B and pin 2 of IC1A and compared with the respective reference. The output of comparator IC1B goes high when U_a exceeds the reference, whereas the output of IC1A goes high when U_a lies below the voltage at pin 3.

The comparators are followed by driver stages, T1 and T2, for the LED display (D1 for "high" and D2 for "low") and also NOR gate IC2A that switches on T3 when the output of both comparators is low, that is, when it is undefined. This state is indicated by D3.

The remaining three gates in IC2 form a monostable. During quiescent operation, U_{CC} is present at the input of inverter IC2C. The output of the inverter is then low, T4 is off and D4 is out. Pin 4 of IC2B is also high, but this state changes when a pulse arrives at pin 5. The output of IC2B then goes low, C2 discharges, the inverter toggles, T4 is switched on, and D4 lights. This state is unstable, however, because C2 recharges via R13. Although the pulse at pin 5 might be very short, the time constant R_{13}/C_2 lengthens it to about 100 ms.

The supply voltage can lie between 5 and 15 V. At 5 V, the circuit draws a current of about 15 mA. The input impedance of the tester is of the order of 330 kΩ.

UNIVERSAL TEST PROBE



ELEKTOR ELECTRONICS *Fig. 67-3*

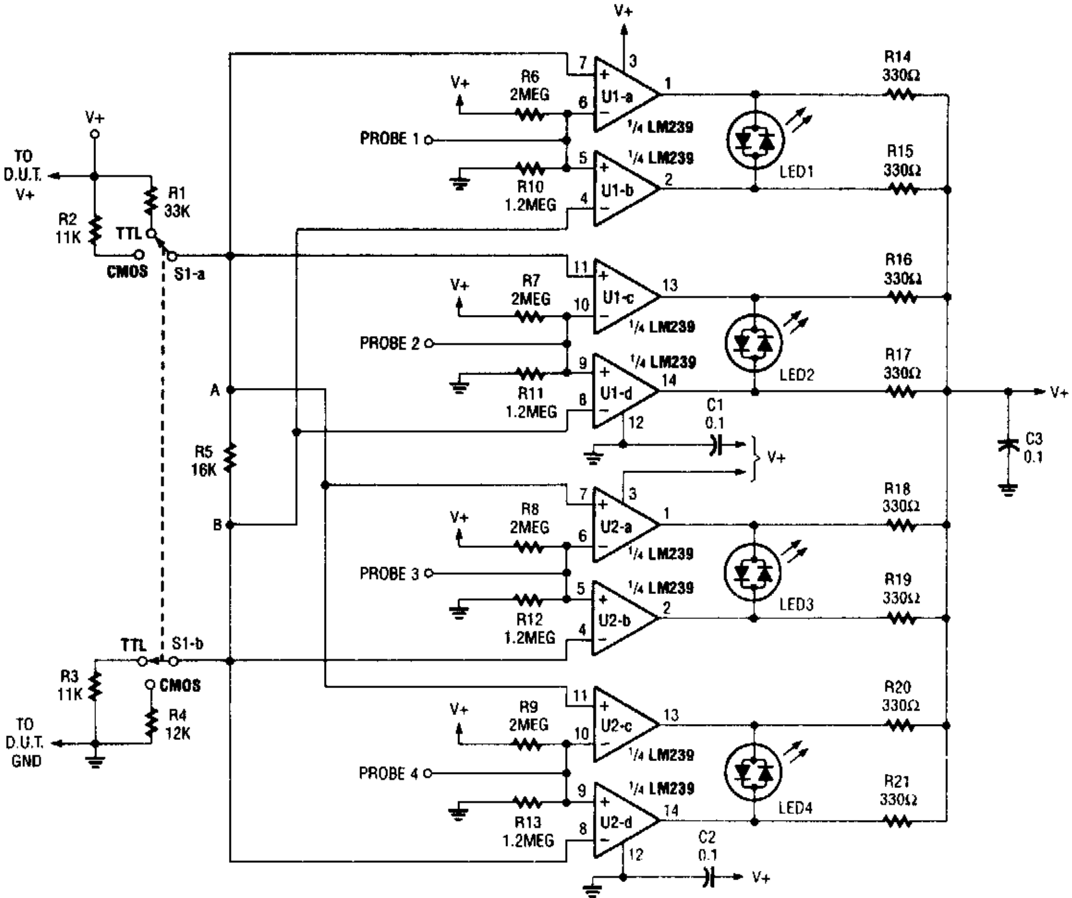
The compact test probe provides rapid "measurement" of voltage levels at digital gates, fuses, diodes, batteries, and others. It does not provide absolute values, but rather it provides a good indication of correct operation or otherwise.

Measurements are carried out with pins A and B. If the potential difference between A (the reference pin) and B is 1.9 to 2.0 V, D2 will light. If the voltage at B is ≤1.4 V higher than that at A, D3 will light. Finally, if the potential at B is ≥11 V, with respect to that at A, D1 will light.

Transistor T5 is used as a zener diode. The probe allows the measurement of alternating voltage. The maximum input voltage is highly dependent on the dissipation allowed in R1. For example, when this resistor is a 0.5-W type, the input voltage can be as high as 200 Vrms.

The current drawn by the circuit depends on the number of lighting LEDs: it is not more than 10 mA at a supply voltage of 3 V. In quiescent operation, the current is so low (about 5 μA) that an on/off switch is not necessary.

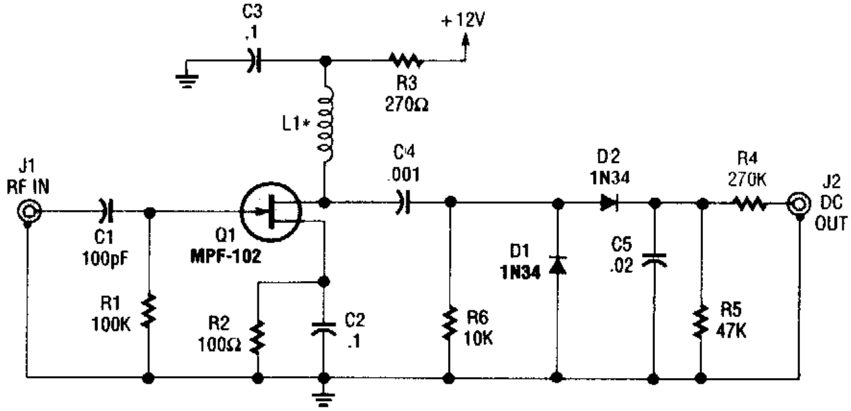
4-WAY LOGIC PROBE



POPULAR ELECTRONICS *Fig. 67-4*

This logic probe has four channels and uses two IF quad comparators to drive four bicolor LEDs. S1 and S1B program the comparator trip levels for TTL and CMOS. R6 through R13 bias the probe inputs to prevent the probe from indicating a HIGH for an open circuit. An open circuit will produce an OFF indication on the LED. The LEDs will indicate one color for high, the other color for low, and intermediate colors for pulsing (assuming a duty cycle between 30 and 70%). The color that corresponds to HIGH or LOW depends on how you connect the LEDs.

ACTIVE RF DETECTOR PROBE

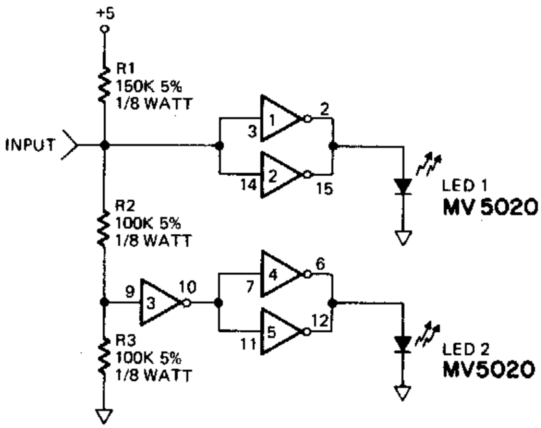


POPULAR ELECTRONICS

Fig. 67-5

An MFP102 FET is used as a wideband amplifier. L1 can be 100 μH for 30 to 100 MHz or 1 000 μH for 2 to 30 MHz. For LF, (less than 3 MHz), use a 2.5-μH RF choke. This probe will work as a relative indicator device, but it can be calibrated over a frequency range, if needed.

SINGLE-IC LOGIC PROBE

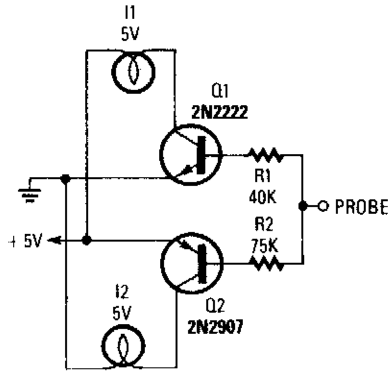


RADIO-ELECTRONICS

Fig. 67-6

This logic probe uses a CD4009 CMOS hex inverter. The characteristic high-input impedance of CMOS gives the advantage of not loading the circuit being tested. Because the output of the inverters is not specified at either a high or low level with a floating input, an input-bias network produces lows at both input inverter pairs if the input is open or at less than 2 V. Resistor R3 holds the input of inverter 3 low which makes the output of inverters 4 and 5 low and will not permit LED 2 to light. At the same time, R1 holds the inputs of inverters 1 and 2 high so that their output is low and LED 1 will not light. If the probe input is touched to a logic low, the output of inverter 3 is held high by R3 and inverter 1 and 2 are brought low, which causes their outputs to go high and turns on LED 1. With no input, both LEDs should be off.

LOGIC PROBE

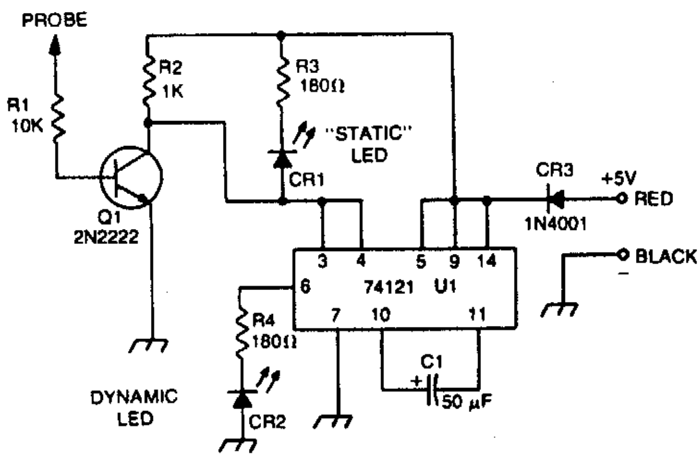


POPULAR ELECTRONICS/HANDS-ON ELECTRONICS **Fig. 67-7**

By connecting this circuit to a logic device that's under power, you can get an indication as to its status. If the circuit is open, neither of the test lamps will light. If the circuit is grounded, the low (or zero) lamp will light. If 3 to 6 V are present, the high-voltage lamp will light.

Other than its application in logic testing, the probe is also convenient for checking supply voltages and grounds. You can select resistors to turn the lamps on at any desired threshold voltage within the component limits.

LOGIC PROBE



POPULAR ELECTRONICS/HANDS-ON ELECTRONICS **Fig. 67-8**

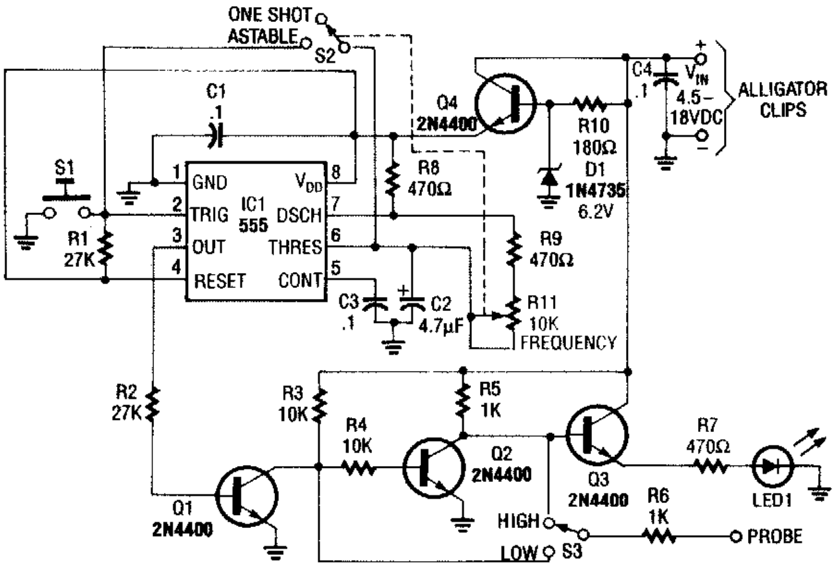
The "static" LED indicates a logic level of 1 when lit. The "dynamic" LED indicates a logic 1 *pulse*. A 4-V, 100-ns pulse will be stretched to about 50 ms so that it can be easily seen.

68

Pulse Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

INEXPENSIVE PULSE GENERATOR FOR LOGIC TROUBLESHOOTING

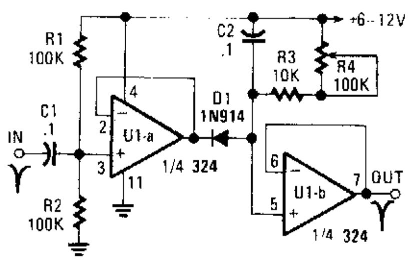


RADIO-ELECTRONICS

Fig. 68-1

Built around a 555 timer IC, this pulse generator can be built into a probe for logic troubleshooting. R11 is frequency controlled and gives a range of about 5 to 200 Hz. C2 can be reduced for higher frequencies. S2 selects one shot or pulse operators. Q1 and Q2 provide a fast rise-time pulse, which acts as a clipper amplifier. Q4 acts as a regulator. The supply can be anything from about 4.5 to 18 Vdc. LED1 is an indicator that shows circuit operation.

NEGATIVE PULSE STRETCHER

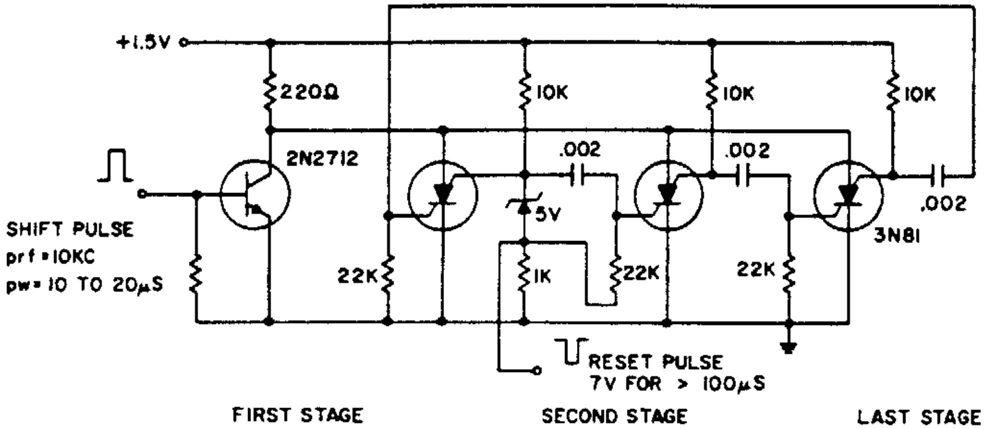


POPULAR ELECTRONICS

Fig. 68-2

U1A acts as an amplifier, which drives D2 and charges C2. U1B acts as a voltage follower. R3 and R4 determine the amount of stretch that the input pulse receives. C2 can be charged to accommodate different pulse rates.

LOW-POWER RING COUNTER (LESS THAN 6 mW)

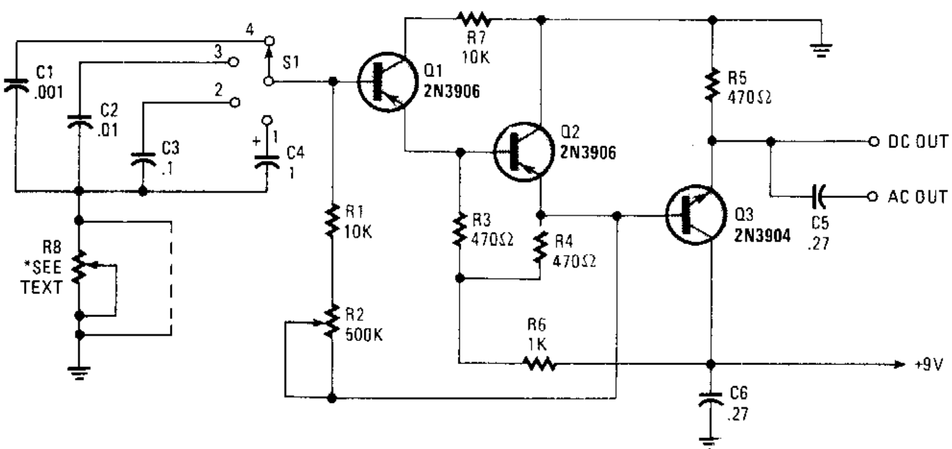


GE

Fig. 68-3

The ring counter operates from 1.0 to 6.0 V and requires only 6 mW at 1.5 V. The reset pulse turns on the first stage with its trailing edge. The maximum shift pulse width increases with voltage and approaches 70 μs for a 6.0-V supply. Minimum pulse width is 10 μs.

TRANSISTOR PULSE GENERATOR

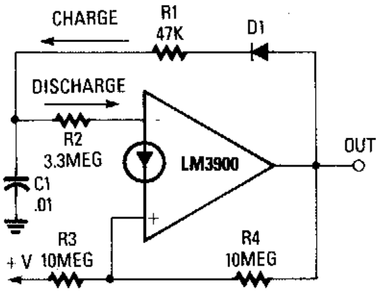


POPULAR ELECTRONICS

Fig. 68-4

Seven-V narrow pulses from 2 Hz to 50 kHz are produced by this circuit. C1 through C4 provide frequency ranges in decode steps. R1 and R2 control the charging time of C1 through C4. R2 is a potentiometer used to set the frequency. R8 controls pulse width. Pulse width varies from 7 μs to 10 ms. Depending on the frequency, R8 can be deleted if it is not needed.

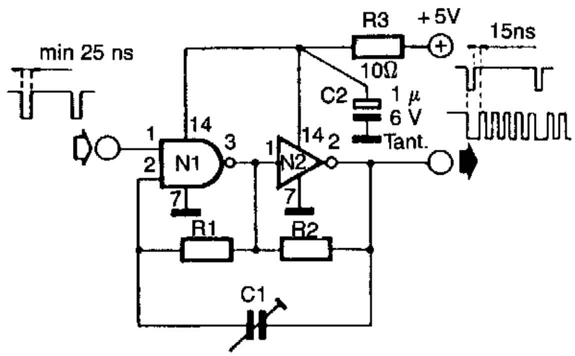
FREE-RUNNING PULSE GENERATOR



RADIO-ELECTRONICS *Fig. 68-5*

C1 alternately charges via R1/D1 and discharges via R2, which produces a duty cycle of about 1:60.

STABLE START-STOP OSCILLATOR

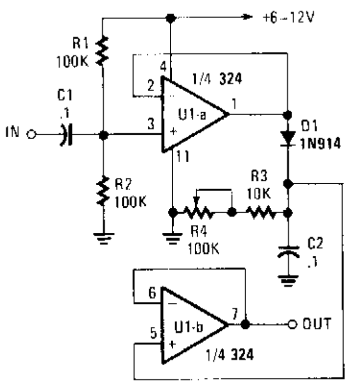


N1 = 1/4 74LS00 R1, R2 = 560 Ω . . . 4k7
N2 = 1/4 74LS04 C1 = 20 . . . 80 p

ELEKTOR ELECTRONICS *Fig. 68-6*

Oscillators that generate a predetermined number of pulses are often required in applications such as video work. This oscillator starts 13 ms after the control signal goes high and stops immediately when the input signal goes low.

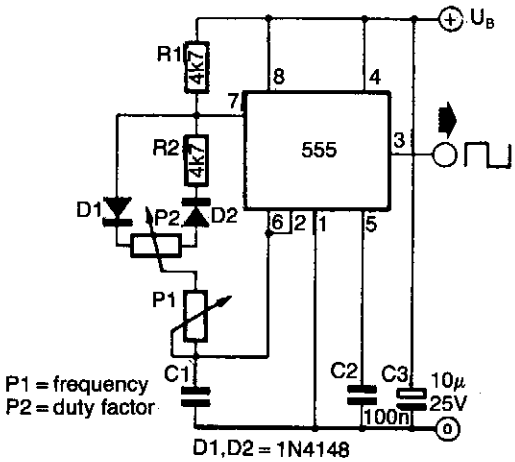
POSITIVE PULSE PULSE STRETCHER



POPULAR ELECTRONICS *Fig. 68-7*

A simple pulse stretcher built with two sections of an op amp uses voltage follower U1A to drive D1 and C2. C2 charges to the peak value of pulse voltage. R3 and R4 determine the discharge time of C2 and therefore the pulse stretching. U1B acts as a voltage follower. Typically this circuit can stretch a pulse by a factor of 50. C2 can be charged to accommodate different pulse rates.

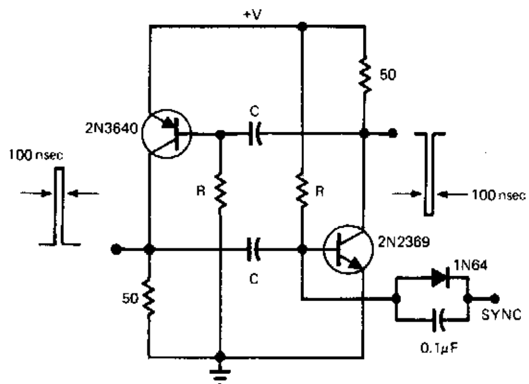
555 PULSE GENERATOR



ELEKTOR ELECTRONICS *Fig. 68-8*

This approach to using a NEC555 timer uses two diodes to set the charge and discharge timer of capacitor, which gives the circuit a variable duty factor.

FAST LOW DUTY-CYCLE PULSE OSCILLATOR

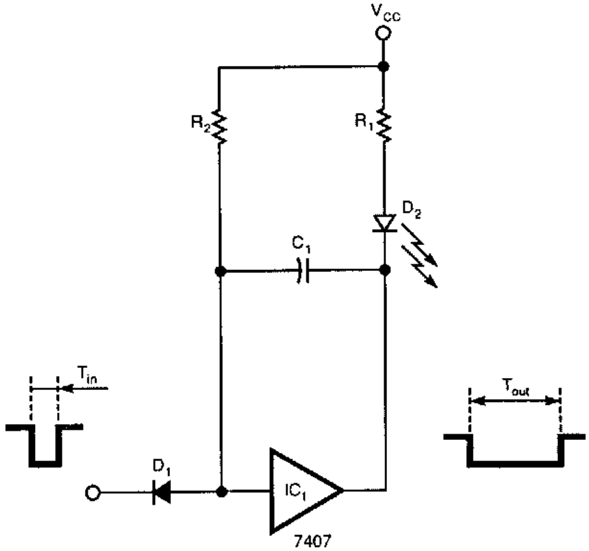


EDN *Fig. 68-9*

This simple and symmetrical free-running generator has a 50-Ω output impedance, a pulse width of 100 ns and complementary outputs that swing essentially from ground to the power-supply voltage. Moreover, it functions with a power supply range from <1 to >15 V and maintains a low voltage and temperature drift while consuming little power.

For oscillation to occur, each transistor must have a gain greater than unity. This restricts the value of R to a range of 1 kΩ to 1 MΩ, because the gain will be less than unity when the transistor is saturated or when beta is low as a result of small collector currents. The two RC timing networks do not have to match because the RC with the longest time constant will determine the frequency of oscillation.

SIMPLE PULSE STRETCHER

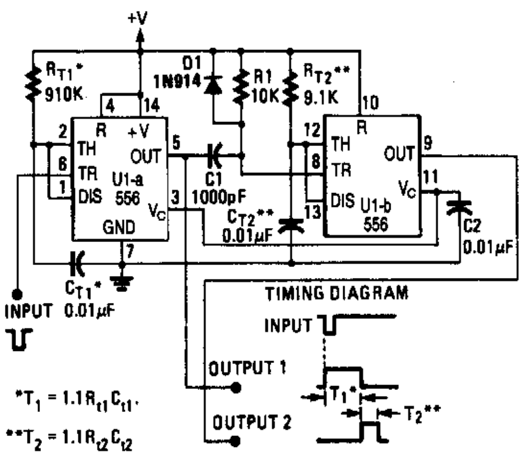


EDN

Fig. 68-10

A single gate (open collector, noninverting) produces a simple one-shot to produce a pulse that stretches equal to the pulse duration, plus the R_1C_1 time constant. R2 is a pull-up resistor to keep the gate's input high while waiting for a pulse.

DELAYED PULSE GENERATOR



HANDS-ON ELECTRONICS (AS SHOWN, T_1 = 10 MS; T_2 = 100µS.)

Fig. 68-11

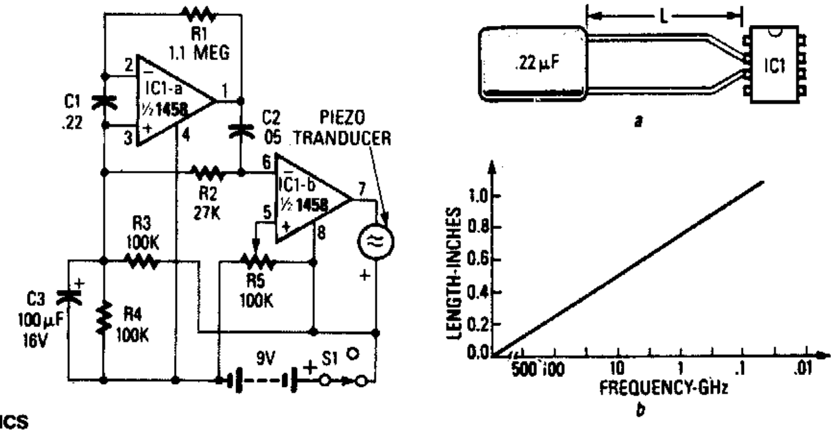
This circuit produces a delayed pulse width of $1.1(R_{t2}C_{t2})$. The delay is given by $T=1.1(R_{t1}C_{t1})$. A 556 dual timer can be used.

69

Radar Detectors

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

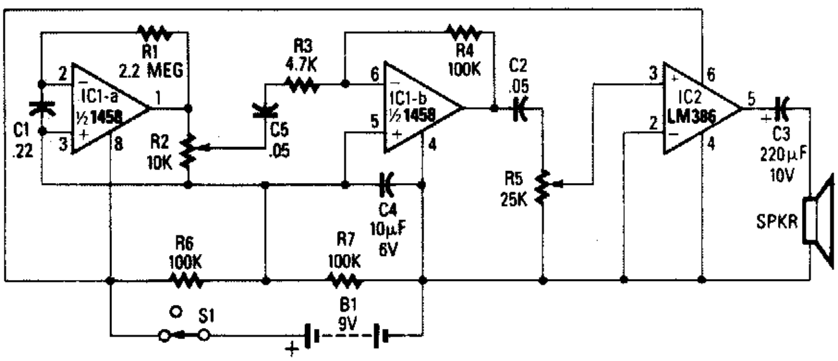
RADAR DETECTOR

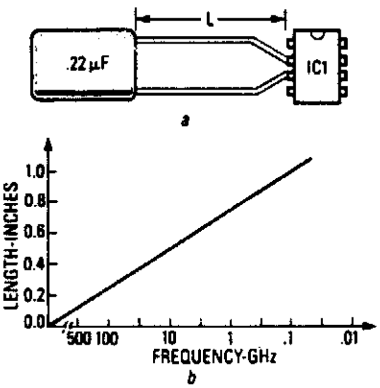


RADIO-ELECTRONICS *Fig. 69-1*

A simple detector of radar signals uses the junctions of the input transistors of an LM1458 op amp as RF detectors. The leads of C1 act as an antenna and should be about 0.4″ long measured from the IC package. Detected audio components are further amplified by IC1A and IC1B, and drive a piezo transducer. Mount the circuit so that incident RF energy will not be blocked.

DELUXE RADAR DETECTOR





This simple radar detector includes an audio amplifier for driving a loud speaker. As in Fig. 69-1, it uses an op amp as a detector of microwave signals.

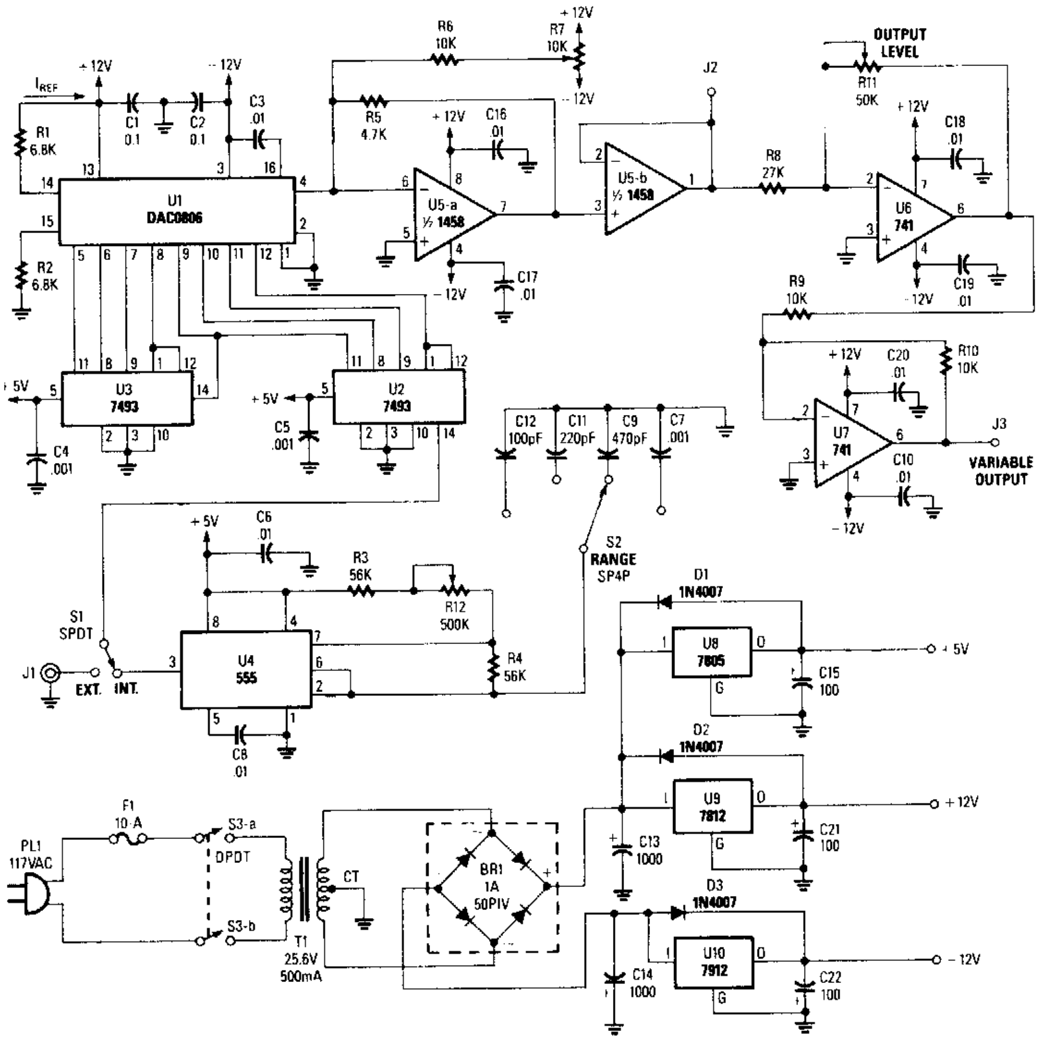
RADIO-ELECTRONICS *Fig. 69-2*

70

Ramp and Staircase Generators

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

DIGITAL SAWTOOTH GENERATOR

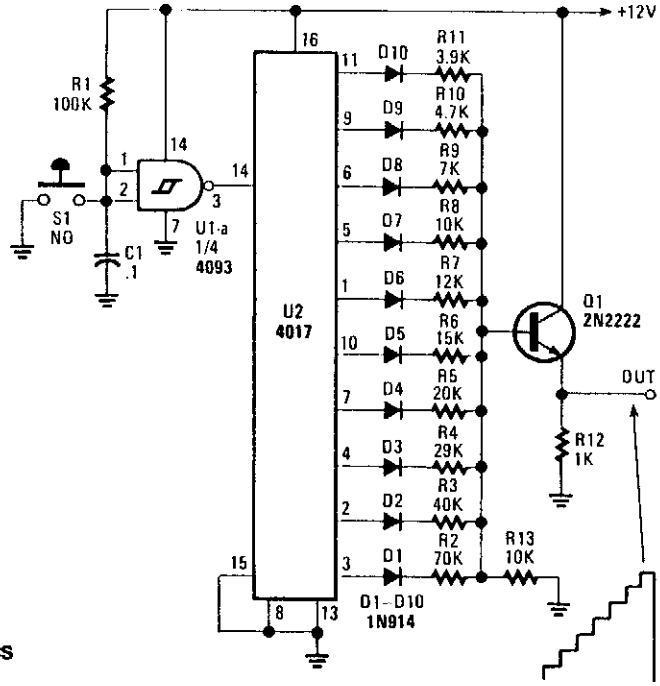


POPULAR ELECTRONICS

Fig. 70-1

This straightforward circuit uses an NE555 timer as an oscillator, a 7493 counter chain and a DAC0806 D/A converter. The counter output feeds the DAC, producing a linearly increasing voltage, which drives an op amp and a dc-level insertion circuit. The clock frequency must be 256 times the desired sawtooth output frequency.

SIMPLE STAIRCASE GENERATOR

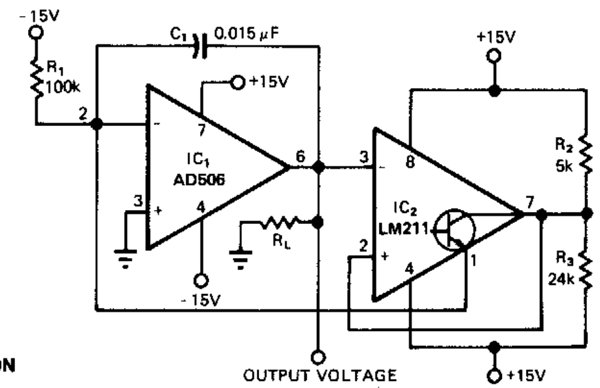


RADIO-ELECTRONICS

Fig. 70-2

U2 is a decade counter /divider. U1 is used as a switch debouncer. For a self-generating system, connect a resistor between pins 2 and 3 of a U1 value that should be between 10 kΩ and several MΩ, depending on desired frequency. C1 can also be varied to change frequency. Also, S1 can be omitted in the self-generating version.

RAMP GENERATOR

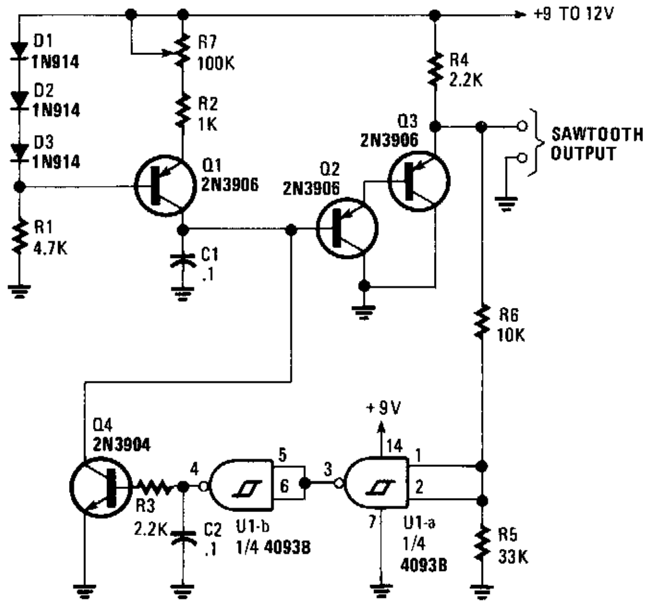


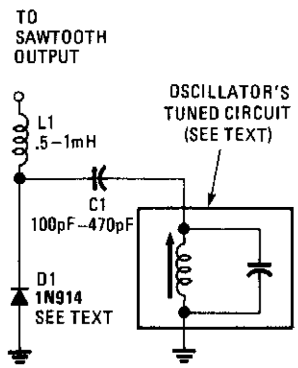
EDN

Fig. 70-3

Providing a 0-to 10-V excursion from 0.4 Hz to 100 kHz, this circuit offers both simplicity and small size. The negative current through R1 produces the ramp's positive slope and causes the output of IC1 to increase linearly toward the +15-V rail.

SAWTOOTH GENERATOR FOR SWEEP GENERATORS



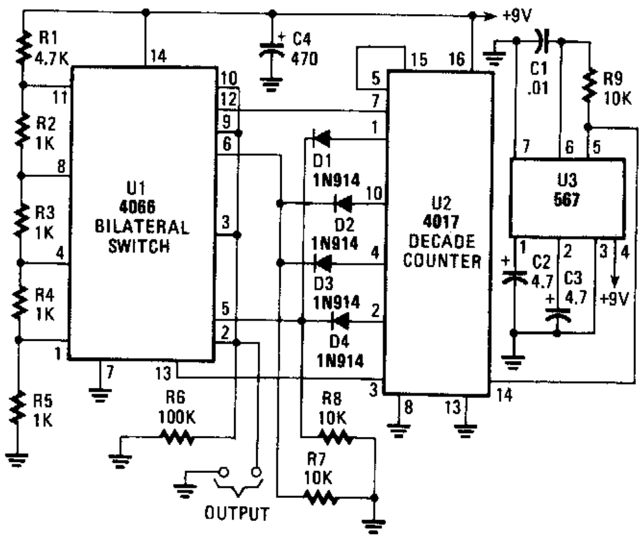


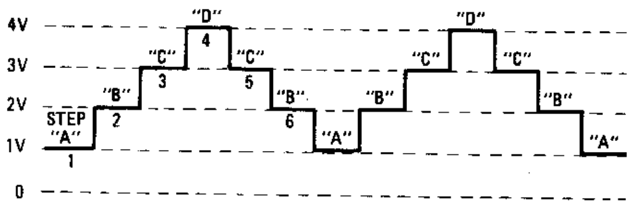
RADIO-ELECTRONICS

Fig. 70-4

This circuit will generate a linear sawtooth between 30 Hz and 3 000 Hz. Q1 is a constant-current source that charges C1 until the output level at Q3 emitter triggers U1A and U1B, which turns on Q4 and discharges C1. The frequency range can be varied by changing the value of C1. This circuit should be good to several tens of kHz.

STEPPED WAVEFORM GENERATOR





The output of the stepped waveform generator is made up of 3-up and 3-down steps in 1-volt increments (much like the output of a digital-to-analog converter). Switch trigging for the 4066 is controlled by a 4017 decade counter/divider (U2), which is clocked by a square-wave generator built around a 567 PLL (phase-locked loop).

POPULAR ELECTRONICS ***Fig. 70-5***

A decode counter (U2) is used to perform sequential switching via a CD 4066. Analog switch is to generate a waveform. The clock is a 567 PLL or other VCO chip.

71

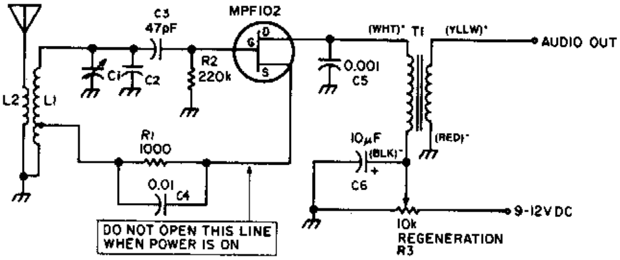
Receivers

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

1-Transistor Regenerative Receiver
3.5- to 10-MHz Simple Superheterodyne Receiver
Simple Low-Frequency Receiver
Reflex Radio Receiver
TRF Radio
Old-Time Radio
Pulse-Frequency Modulated Receiver
Shortwave Receiver
Simple AM Radio

1-TRANSISTOR REGENERATIVE RECEIVER

RECEIVER



T1 = ANY AUDIO TRANSFORMER AS LONG AS PRIMARY IS GREATER THAN 1kΩ

T50-2 $N = 100\sqrt{L/50}$
N = NUMBER OF TURNS
L = INDUCTANCE IN μH

*COLOR CODED CONNECTIONS FOR 1:1 AUDIO TRANSFORMER

L1 & C1 = RESONANT FREQUENCY
USE FORMULA

$$f = \frac{10^6}{2\pi\sqrt{LC}}$$

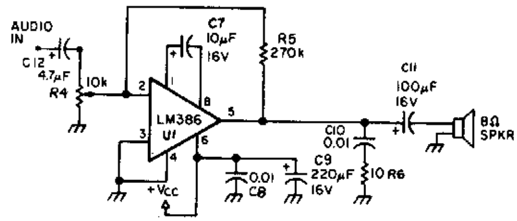
TAP 25% FROM COLD END

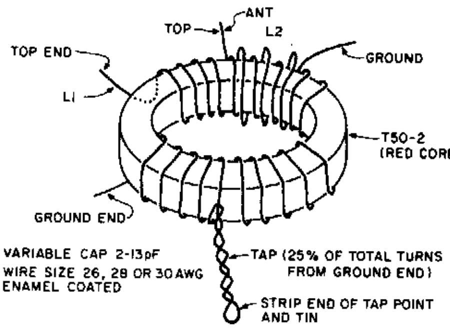
L2 = 3 TURNS AROUND L1

COIL FORM CAN BE ANYTHING
EVEN T50-2 TOROID

$$f = \frac{10^6}{2\pi\sqrt{LC}} \qquad L = \frac{10^6}{39.5 \cdot f^2 \cdot C} \qquad C = \frac{10^6}{39.5 \cdot f^2 \cdot L}$$

f = FREQUENCY IN MHz
L = INDUCTANCE IN μH
C = CAPACITANCE IN pF





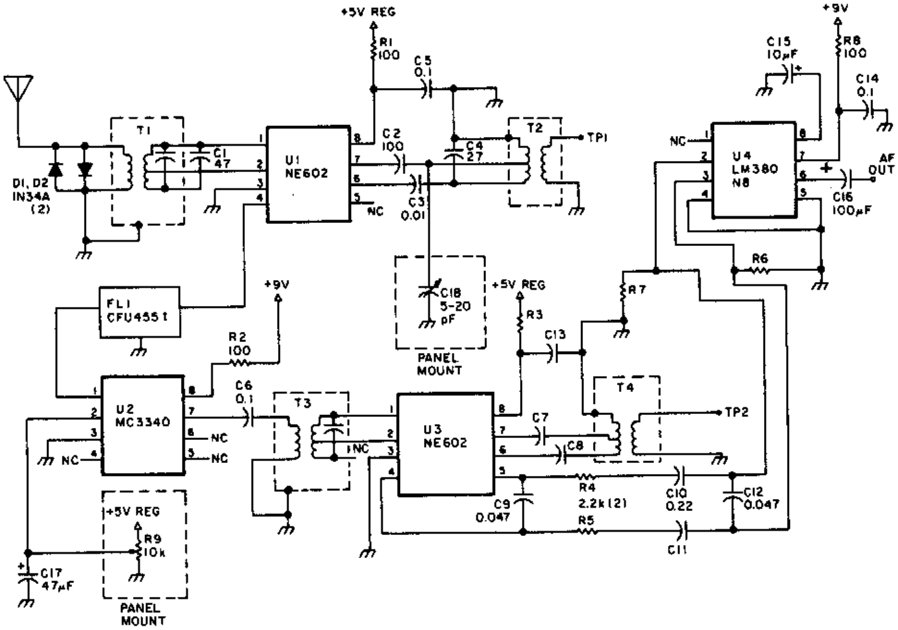
Using an MPF102 FET, this one-transistor radio receiver is suitable for experimental use. It can tune 4 to 26 MHz. For speaker operation, a simple IC amplifier can be used.

Frequency (MHz)	C2 (pF)	C18, C19 (pF)	C23 (pF)	L1, L2 (Ant) (# of turns on T-37-2 core)	L3 (Osc)
5	100	120	68	5, 41	45
6	100	120	68	4, 30	34
7	82	100	47	4, 26	29
8	82	100	47	3, 22	24
10	82	100	47	3, 17	19
12	82	100	47	2, 15	17
14	68	82	33	2, 14	15
15	68	82	33	2, 13	14

73 AMATEUR RADIO

Fig. 71-1

3.5- TO 10-MHz SIMPLE SUPERHETERODYNE RECEIVER



SupeRX Parts List

Part	Value	Type
C1	47 pF	silver-mica or polystyrene
C2	100 pF	silver-mica or polystyrene
C3	0.01µF	polystyrene or monolithic
C4	27 pF	silver-mica or polystyrene
C5	0.1µF	ceramic disc or monolithic
C6	0.1µF	polystyrene or monolithic
C7	0.022µF	polystyrene or monolithic
C8	0.1µF	polystyrene or monolithic
C9	0.047µF	monolithic
C10	0.22µF	monolithic
	0.15µF*	(alternate)
C11	0.22µF	monolithic
	0.15µF*	(alternate)
C12	0.047µF	monolithic
C13, C14	0.1µF	ceramic disc or monolithic
C15	10µF	electrolytic, 16V
C16	100µF	electrolytic, 16V
C17	47µF	electrolytic, 16V
C18	5–20 pF	panel mounted tuning capacitor
C19, C20	220µF	electrolytic, 16V
D1, D2	1N34A	germanium diode or equivalent
R1, R2, R3, R8	100 ohm, ¼W	carbon composition
R4, R5	2.2k, ¼W	carbon composition.
	1.5k	(alternate)

73 AMATEUR RADIO

Fig. 71-2

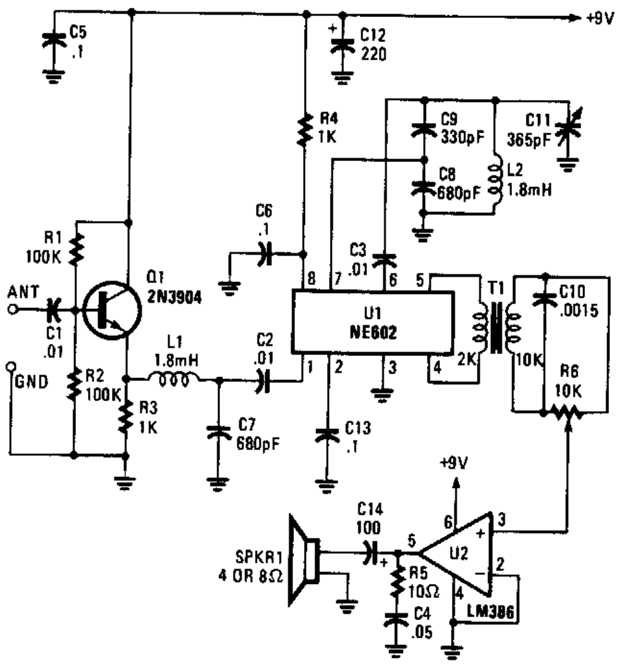
R6, R7	10k, ¼W	carbon composition	
R9	10k	potentiometer	
T1, T2	10.7 MHz	microminiature (7mm) IF transformer, green core	Mouser PN 42IF223
T3, T4	455 kHz	microminiature (7mm) IF transformer, black core	Mouser PN 42IF203
U1, U3	NE602	double-balanced mixer	
U2	MC3340	variable attenuator	
U4	LM380N-8	audio amplifier	
U5	78L05	100 mA miniature +5V regulator	

* C10 and C11 can range from 0.1 to 0.22µf. Valves greater than 0.33 cause distortion.

Other: Printed stripboard, DSE PN H5614 or equivalent, cabinet, plastic stick-on feet, 4-40 hardware, etc.

In this circuit, U1 is a frequency converter that feeds the 455-kHz IF stage U2 and detector U3. U4 is the audio output stage. R9 is a gain control that varies the gain of U2. Coil data is given in the part list.

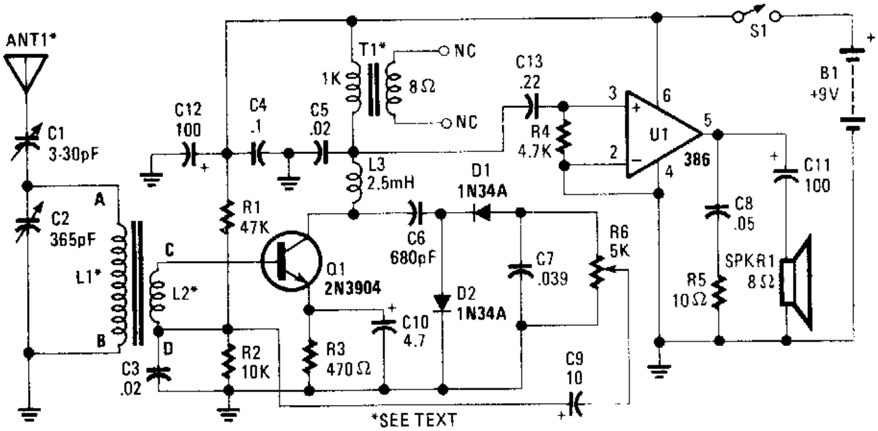
SIMPLE LOW-FREQUENCY RECEIVER



POPULAR ELECTRONICS *Fig. 71-3*

Using an NE602 heterodyne detector and U1 as an RF amplifier, this receiver tunes the middle portion of the low-frequency spectrum from 150 to 250 kHz. U2 is a loudspeaker amplifier.

REFLEX RADIO RECEIVER



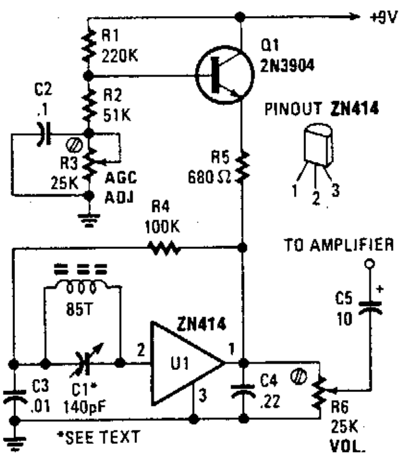
POPULAR ELECTRONICS

Fig. 71-4

The RF signal is passed from the antenna through C1 to the tuned circuit made up of L1 and C2. One end of L2 feeds the RF signal to the base of Q1 for amplification and the other end ties to the junction of R1 and R2 to supply bias to the transistor. A 0.02-μF capacitor, C3, places the "D" end of L1 at RF ground.

The amplified RF signal is fed through C6 to a two-diode doubler/rectifier circuit and then on to the volume control, R6. The wiper of R6 feeds the detected audio signal through C9 to the junction of R1, R2, and the "D" end of L2. The "D" end of L2 is at RF ground, but not AF ground, allowing the AF signal to be passed through L2 to the base of Q1 for amplification. The junction of the 2.5-mH choke and T1 is placed at RF ground through C5. The amplified audio is fed from this junction to the input of the 386 audio amplifier, U1, to drive the 4″ 8-Ω speaker. The single transistor has performed a dual duty by amplifying the RF and AM signals at the same time.

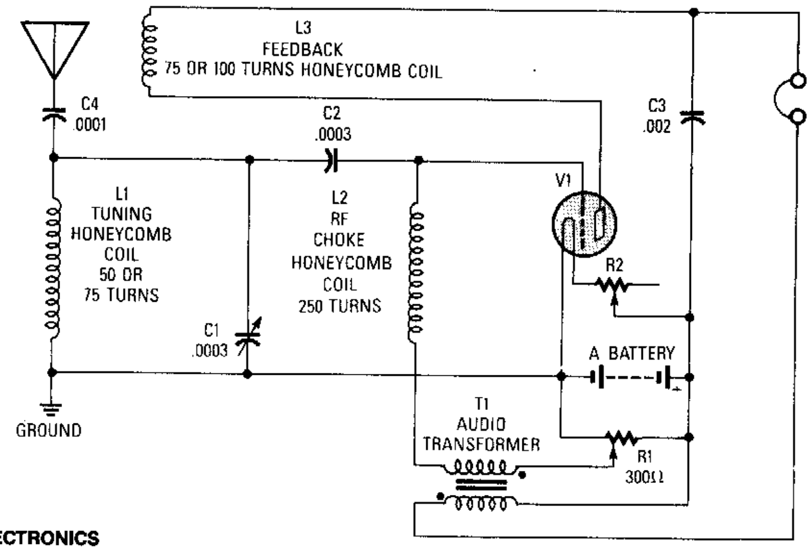
TRF RADIO



POPULAR ELECTRONICS *Fig. 71-5*

This simple AM radio uses a Ferranti ZN414 IC and runs from a 9-V supply. A 5/16″-diameter ferrite rod serves as an antenna, and uses about 85 turns of #28 enamelled wire.

OLD-TIME RADIO

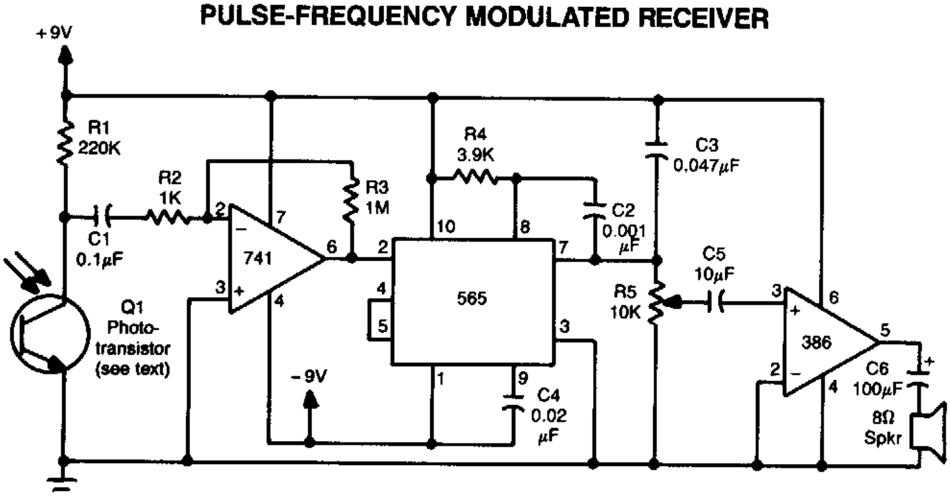


POPULAR ELECTRONICS

Fig. 71-6

This circuit was used in the early days of radio to receive signals. Almost any battery-operated triode, such as a type 30, can be used. "A" battery is 3 V, R2 is a 100-Ω rheostat. Coils are typically 2" to 3" diameter honeycomb wound.

PULSE-FREQUENCY MODULATED RECEIVER

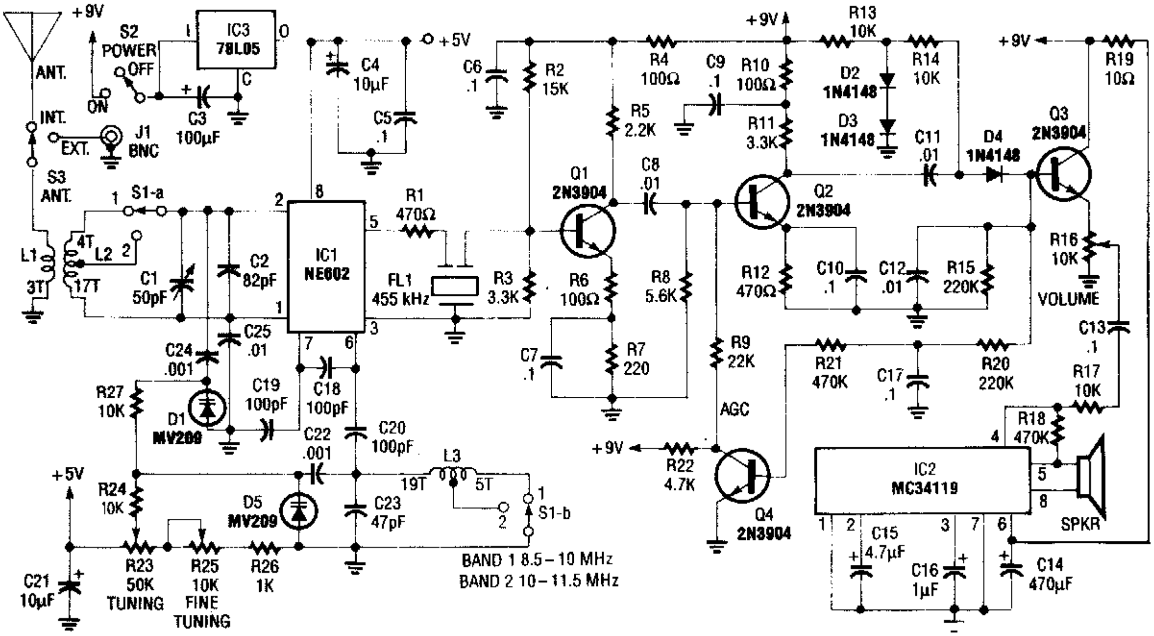


POPULAR ELECTRONICS

Fig. 71-7

This receiver uses an IR-sensitive phototransistor (Clairex, HP, etc.) mounted in a light-tight enclosure with an aperture for the incoming IR beam. An optical system can be used with this receiver for increased range. A 741 amplifies the pulsed IR signal and a 565 PLL FM demodulator recovers the audio, which drives an LM386 audio amplifier and speaker.

SHORT WAVE RECEIVER



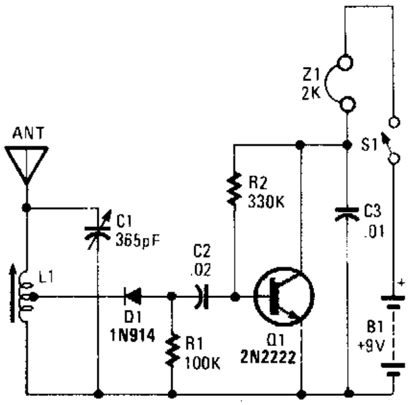
TABLE

Frequency (MHz)	C2 (pF)	C18, C19 (pF)	C23 (pF)	L1, L2 (Ant) (# of turns on T-37-2 core)	L3 (Osc)
5	100	120	68	5, 41	45
6	100	120	68	4, 30	34
7	82	100	47	4, 26	29
8	82	100	47	3, 22	24
10	82	100	47	3, 17	19
12	82	100	47	2, 15	17
14	68	82	33	2, 14	15
15	68	82	33	2, 13	14

RADIO-ELECTRONICS *Fig. 71-8*

Using a Signetics NE602 in a varactor-tuned front end, the circuit of a shortwave receiver can be very simple and yet give high performance. This circuit also uses a ceramic filter as a sensitivity-determining device, two IF stages, AGC, and an audio amplifier. It has a sensitivity of under 1μV. The table shows coil data for the frequencies from 5 to 16 MHz. The values C_{18}, C_{19}, and C_{23} depend on the frequency range chosen.

SIMPLE AM RADIO



POPULAR ELECTRONICS *Fig. 71-9*

An AM radio can be built of a simple diode detector and an audio amplifier. A random length of wire always serves as an antenna. L1 is an adjustable ferrite loopstick of the type used in transistor radios.

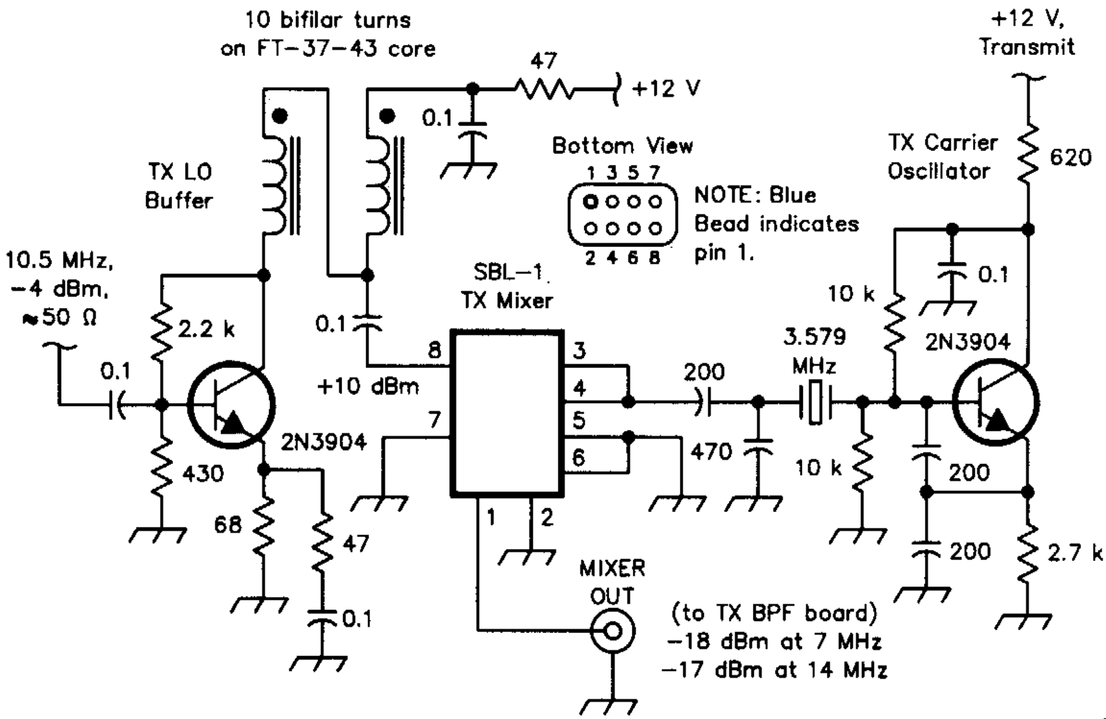
72

Receiving Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 676. The figure number in the box of each circuit correlates to the entry in the Sources section.

HF Transceiver Mixer
AGC System For CA3028 IF Amplifier
Receiver IF Amplifier
30-MHz IF Preamp
Carrier-Operated Relay

HF TRANSCEIVER MIXER

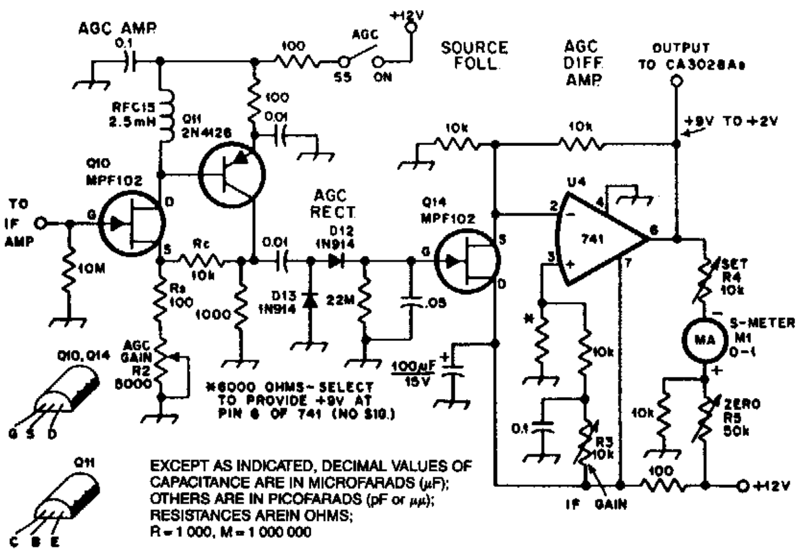


QST

Fig. 72-1

The transceiver mixer and carrier oscillator in the band-imaging (7- and 14-MHz) CW transceiver. Careful selection of drive levels, and use of a spectrally clean carrier oscillator, assure low spurious-signal content in the transmitter output. This transceiver mixer should prove useful in HF and VHF CW or SSB applications. A Mini-Circuits SBL1 low-cost mixer is used with a 3.579-MHz crystal oscillator that uses a low-cost TV color-burst crystal. By paying careful attention to drive leads, good performance and low spurious content can be obtained.

AGC SYSTEM FOR CA3028 IF AMPLIFIER

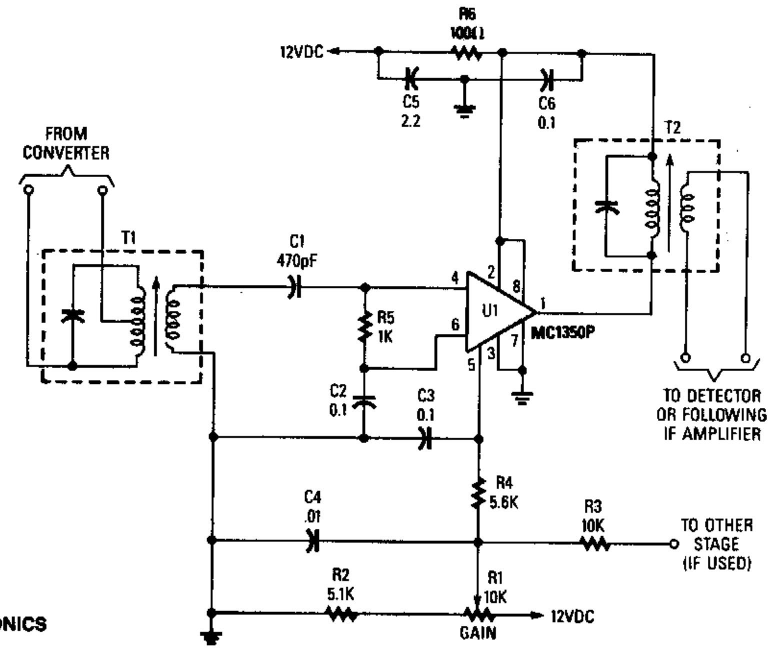


ARRL HANDBOOK

Fig. 72-2

An MPF102 amplifier feeds IF signals to a 2N4126. A potentiometer in the MPF102 source acts as a gain control. This voltage is rectified by an 1N914 doubling detector, and drives a 741 op amp via a source follower (Q14). S-meter and IF-gain controls are provided.

RECEIVER IF AMPLIFIER

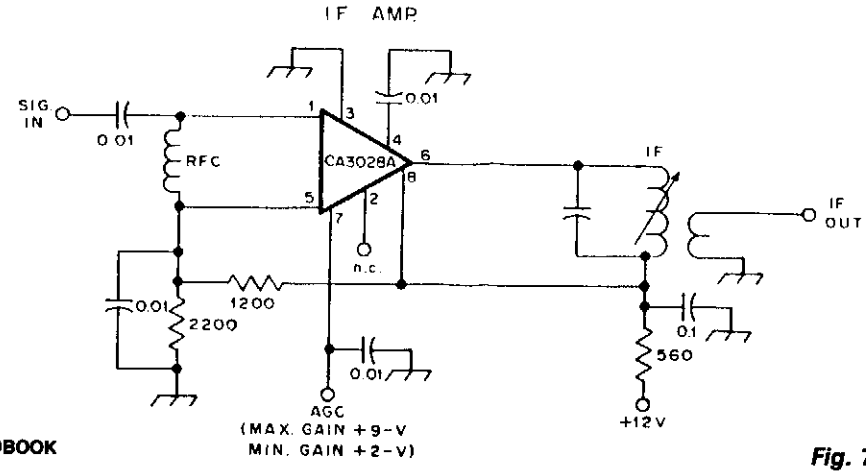


POPULAR ELECTRONICS

Fig. 72-3

T1 is tuned to converter-output frequency U1 to provide 45-to-50-dB gain, depending on the design of T1 and T2. C2, C3, C4, C5, and C6 are bypass capacitors. R5 is a bias resistor. Gain is set by R1, which controls the voltage on pin 5 of U1. T1 and T2 should provide source and load impedance of 1-kΩ and 3- to 10-kΩ, respectively. R3 supplies dc bias to other stages, if required.

IF AMPLIFIER

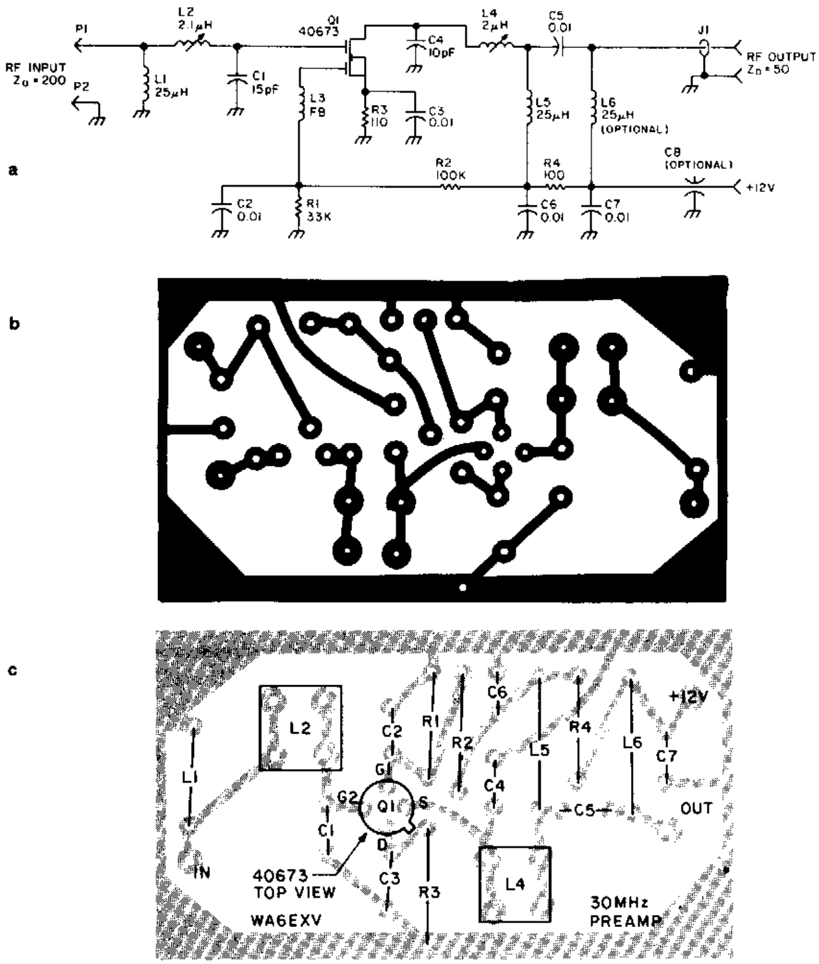


ARRL HANDBOOK

Fig. 72-4

Using a CA3028A, this circuit is useful up to about 15 or 20 MHz.

30-MHz IF PREAMP

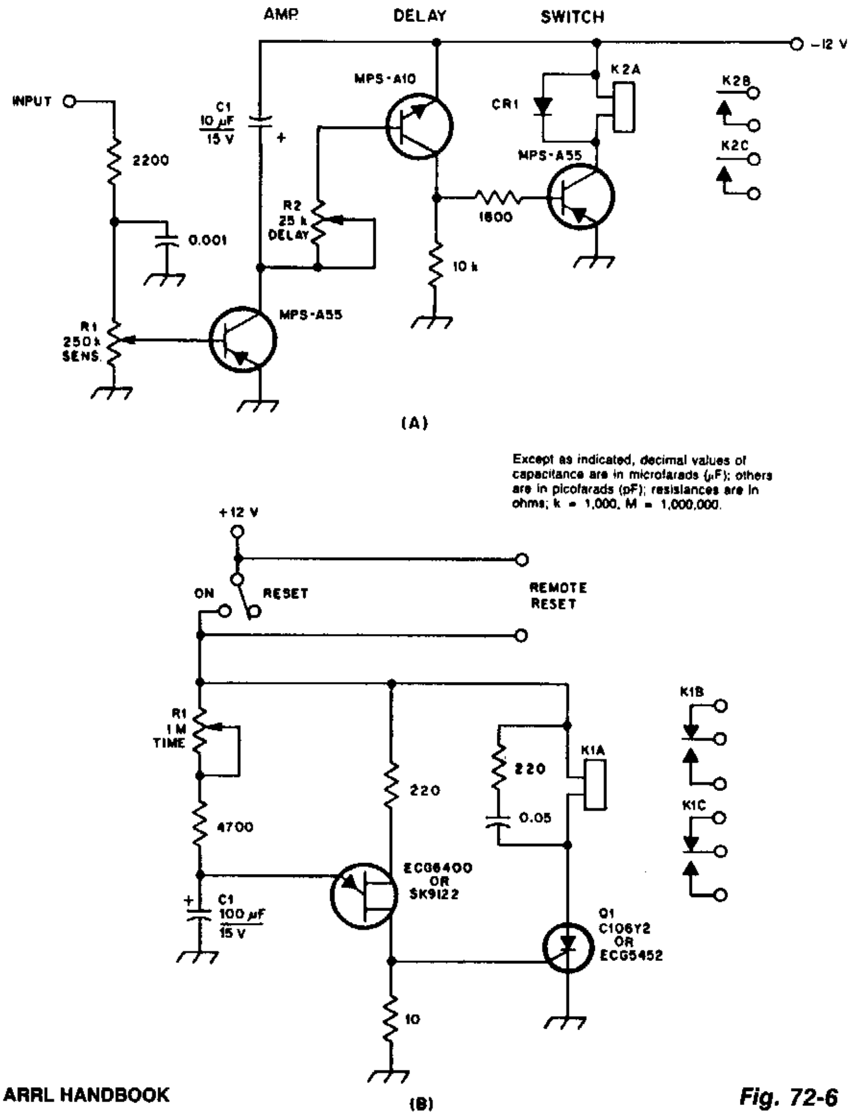


73 AMATEUR RADIO

Fig. 72-5

This preamp for 30 MHz is useful for IF applications used in microwave work, etc. A 40673 MOSFET is used and typically the gain at 30 MHz will be 20 to 25 dB.

CARRIER-OPERATED RELAY



ARRL HANDBOOK *Fig. 72-6*

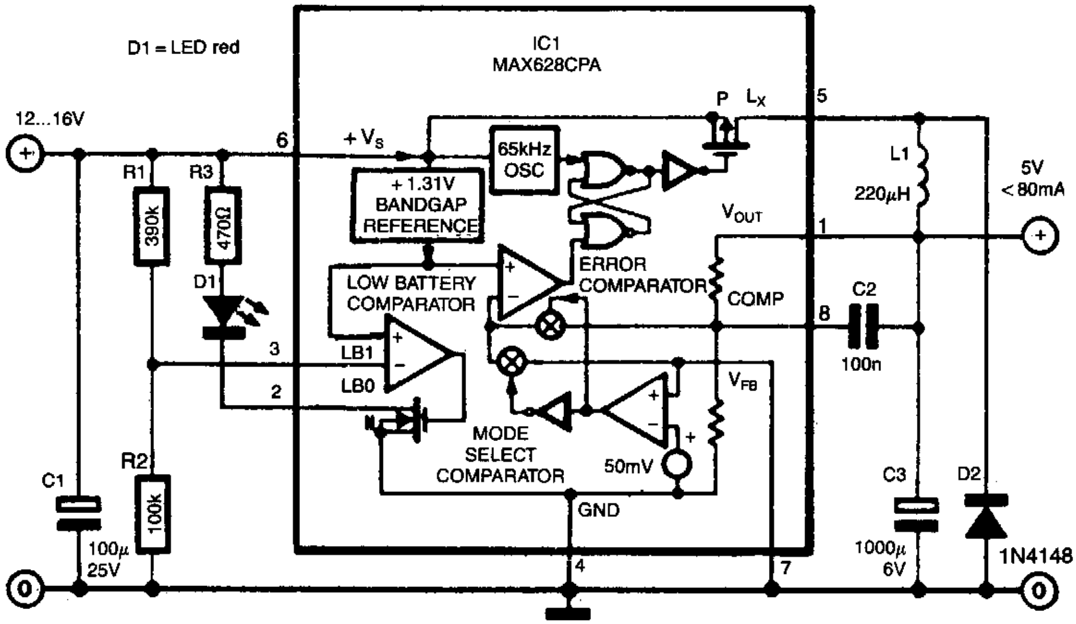
A shows a COR/CAS circuit for repeater use. CR1 is a silicon diode. K2 may be any relay with a 12-V coil (a long-life reed relay is best). R2 sets the length of time that K2 remains closed after the input voltage disappears (hang time). B shows a timer circuit. Values shown for R1 and C1 should provide timing up to four minutes or so. C1 should be a low-leakage capacitor; Q1 is a silicon-controlled rectifier, ECG-5452 or equivalent. K1 may be any miniature relay with a 12-volt coil. The timer is reset when the supply voltage is momentarily interrupted. The switch must be in the RESET position for the remote reset to work. This circuit operates from the detector output of a receiver. A delay circuit is included so that the relay stays closed for a time period after the carrier output from the receiver disappears.

73

Regulators (Fixed)

The sources of the following circuits are contained in the Sources section, which begins on page 676. The figure number in the box of each circuit correlates to the entrv in the Sources section.

SWITCH-MODE VOLTAGE REGULATOR



ELEKTOR ELECTRONICS

Fig. 73-1

Switch-mode power supplies offer the benefit of a much greater efficiency than obtainable with a traditional power supply. The switch-mode regulator presented here has an efficiency of around 85%.

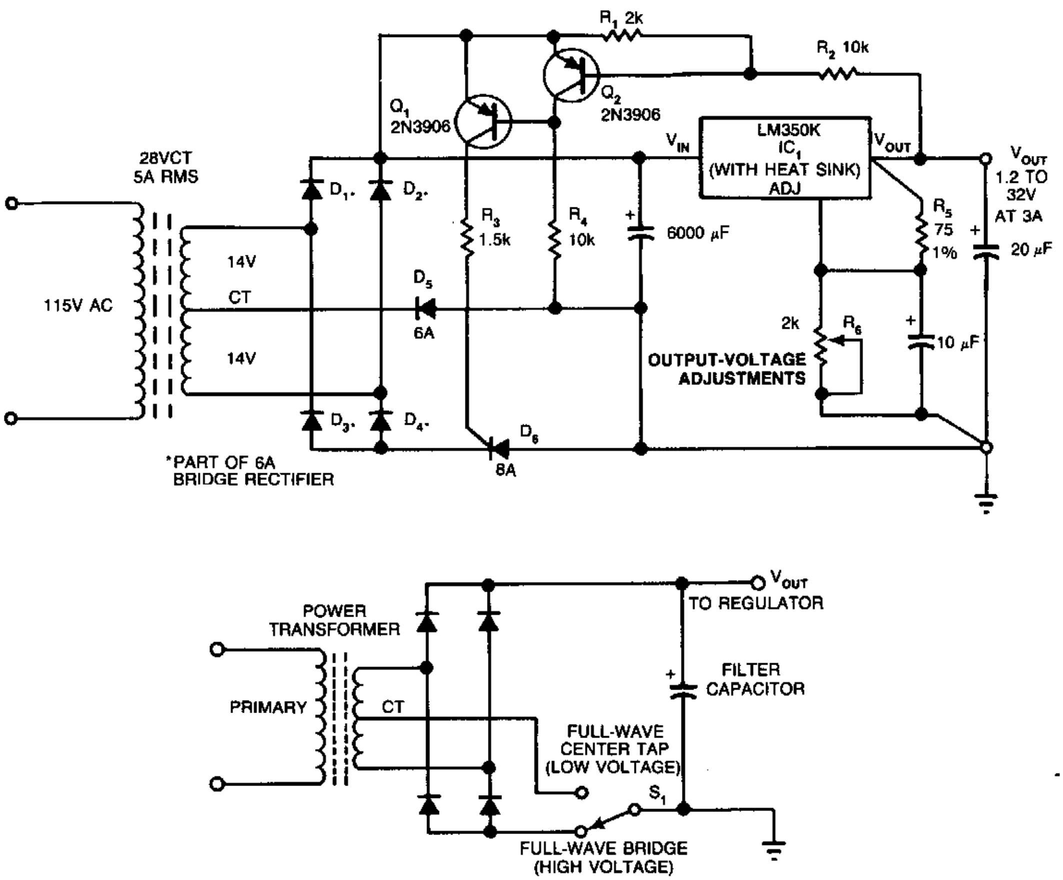
An input voltage of 12 to 16 Vdc is converted into a direct voltage of exactly 5 V. The use of a MAX638CPA enables the design and construction of the regulator to be kept fairly simple: only nine additional components are needed to complete the circuit.

Resistors R1 and R2 are used to indicate when the battery voltage becomes low: as soon as the voltage on pin 3 becomes lower than 1.3 V, D1 lights. With values as shown for the potential divider, this corresponds to the supply voltage getting lower than about 6.5 V. The output of the IC is shunted by a simple LC filter formed by L1, C3 and D2.

The oscillator on board the IC generates a clock frequency of around 65 kHz and drives the output transistor via two NOR gates. The built-in error detector, the "battery low" indicator or the voltage comparator can block the clock frequency, which causes the transistor to switch off.

The IC compares the output voltage of 5 V with a built-in reference (FET). Depending on the load, the FET will be switched on for longer or shorter periods. The maximum current through the FET is 375 mA, which corresponds with a maximum output current of 80 mA.

SWITCHING IMPROVES REGULATOR EFFICIENCY

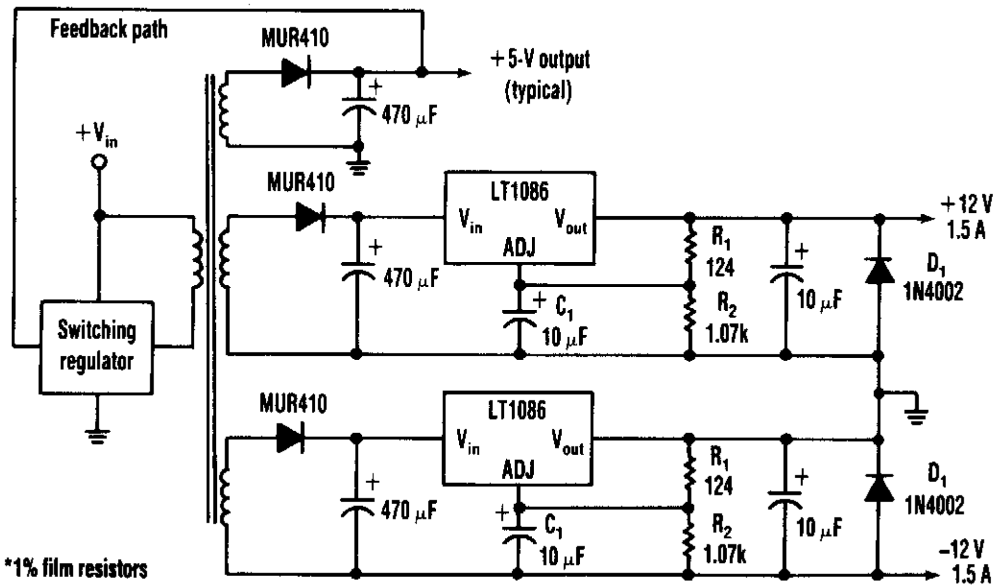


EDN

Fig. 73-2

In this circuit, a full-wave bridge is switched to a full-wave center tap to reduce regulator dissipation. SCR D6 switches between configurations. When D6 is off, the circuit is an FWCT rectifier using D1, D2, and D5. It applies 17 V plus ripple to the regulator input. The drop across the regulator supplies base drive to Q2. If Q2 is on, Q1 is off, and D6 is off. If the regulator voltage drops below about 3 V, Q2 turns off, and turns Q1 on, which turns on D6. This changes the circuit to an FW bridge using D1 through D4.

EFFICIENT NEGATIVE VOLTAGE REGULATION



ELECTRONIC DESIGN

Fig. 73-3

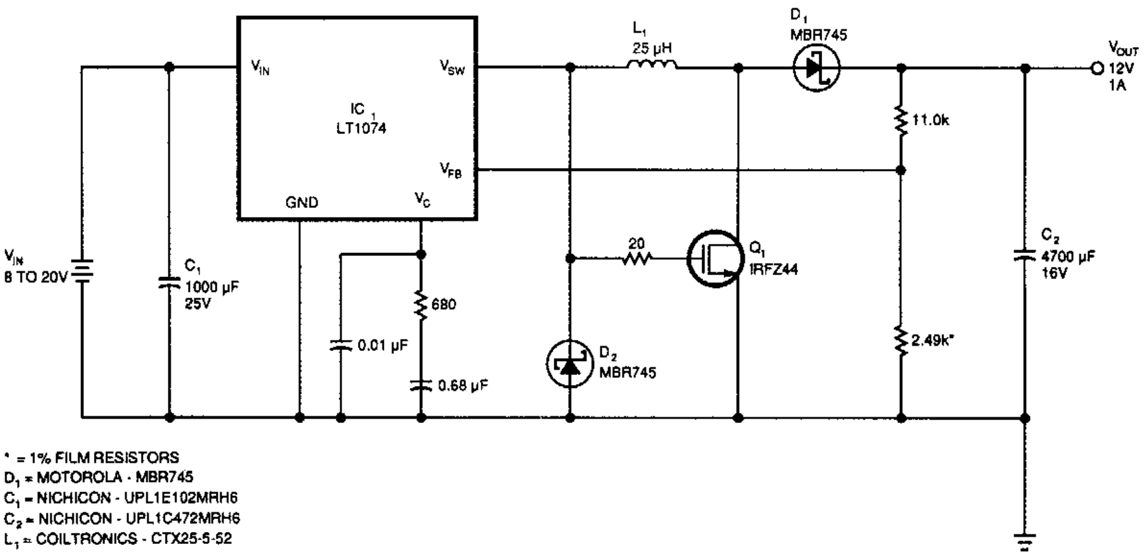
Many applications require highly efficient negative-voltage post regulators with low dropout voltage in switch-mode supplies.

A way to provide good negative-voltage regulation is with a low-dropout positive-voltage regulator operating from a well-isolated secondary winding of the switch-mode transformer. The technique works with any positive-voltage regulator, although highest efficiency occurs with low-dropout types.

In the circuit, two programming resistors, R1 and R2, set the output voltage to 12 V, and the LT1086s servo the voltage between the output and its adjusting (ADJ) terminals to 1.25 V. Capacitor C1 improves ripple rejection, and protection diode D1 eliminates common-load problems.

Because a secondary winding is galvanically isolated, a regulator's 12-V output can be referenced to ground. Therefore, in the case of a negative-voltage output, the positive-voltage terminal of the regulator connects to ground, and the −12-V output comes off the anode of D1. The V_{in} terminal floats at 1.5 V or more above ground. This arrangement is the equivalent of connecting the positive terminal of a battery to ground and taking the output from the negative terminal.

HIGH- OR LOW-INPUT REGULATOR



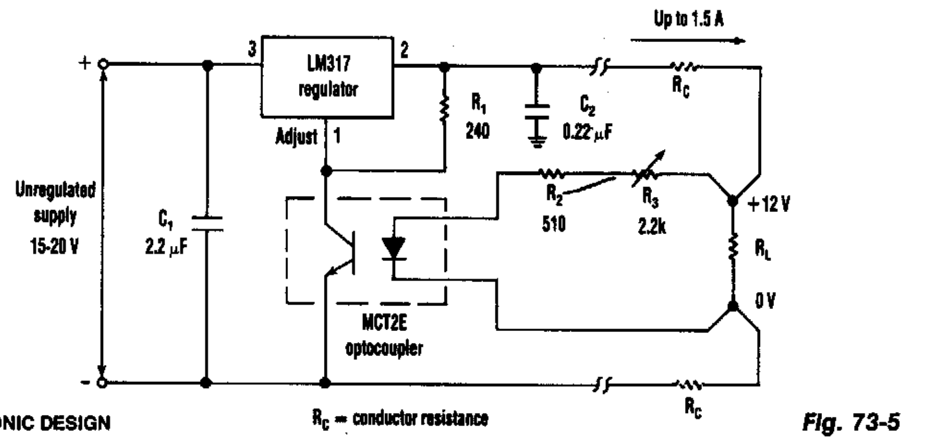
EDN

Fig. 73-4

This regulator provides 12 V at 1 A out with an input voltage of 8 to 20 V. Output voltage can be changed by charging the 11-kΩ and 2.49-kΩ resistors to provide 2.21 V at the V_{FB} pin of IC1, if desired. If you need to handle a higher input voltage, make sure to clamp the gate of Q1 below its 20-V max. rating.

Efficiency can exceed 70% for output currents greater than 0.5 A; above 15-V input voltage, more than 2 A of output current can be obtained.

LM317 REGULATOR SENSING

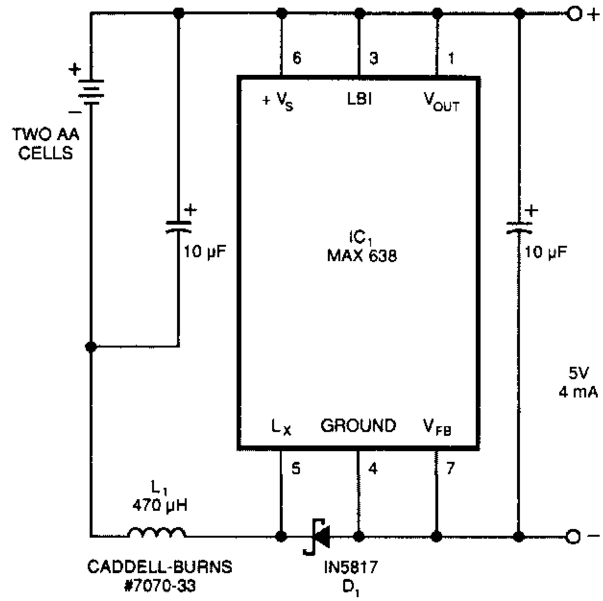


ELECTRONIC DESIGN

Fig. 73-5

The optocoupler (as shown) provides load sensing for a 3-terminal regulator, such as the LM317 series. R1 sets a current of 5 mA through the optocoupler transistor and R3 is adjusted for 12 V across the load.

COMMON HOT-LEAD REGULATOR

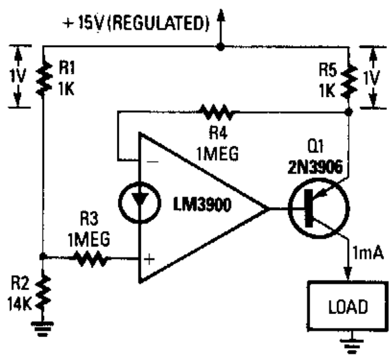


EDN

Fig. 73-6

This circuit derives 5 Vdc from 2-AA cells—even at their end-life voltages of 1.05 V, and is approximately 80% efficient, providing 5 V at 4 mA from 2.1 V at 11 mA. IC1 is manufactured by Maxim Integrated Products, Inc.

FIXED-CURRENT REGULATOR



RADIO ELECTRONICS

Fig. 73-7

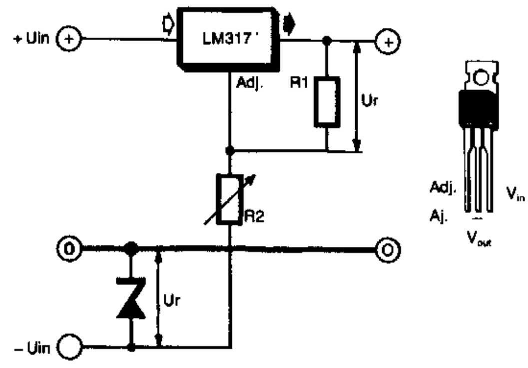
This fixed 1-mA current source delivers a fixed current to a load connected between Q1's collector and ground; the load can be anywhere in the range from 0 Ω to 14 Ω. The circuit is powered from a regulated 15-V supply, and the R1/R2 voltage divider applies a 14-V reference to R3. The op amp's output automatically adjusts to provide an identical voltage at the junction of R4 and R5. That produces 1 V across R5, resulting in an R5 current of 1 mA. Because that current is derived from Q1's emitter, and the emitter and collector currents of a transistor are almost identical, the circuit provides a fixed-current source. The output current can be doubled by halving the value of R5.

74

Regulators (Variable)

The sources of the following circuits are contained in the Sources section, which begins on page 676. The figure number in the box of each circuit correlates to the entry in the Sources section.

REGULATOR CIRCUIT



ELEKTOR ELECTRONICS

Fig. 74-1(a)

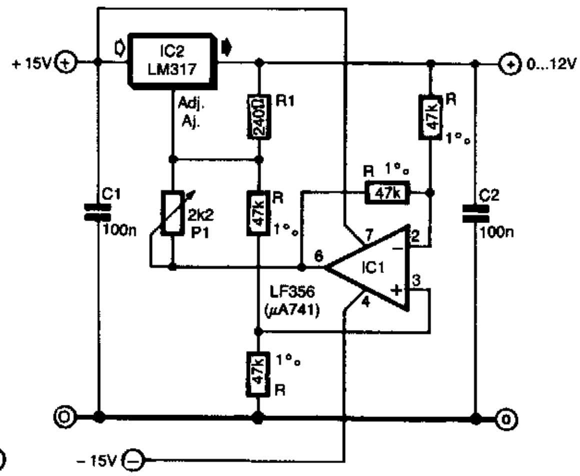


Fig. 74-1(c)

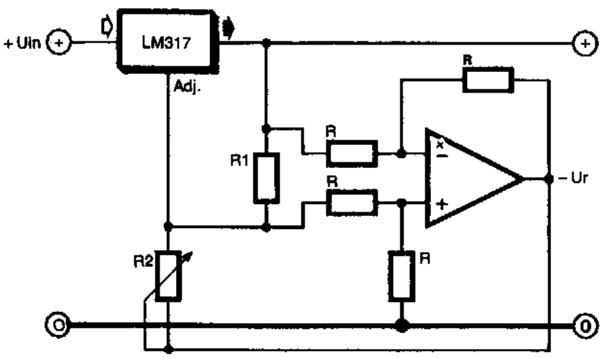


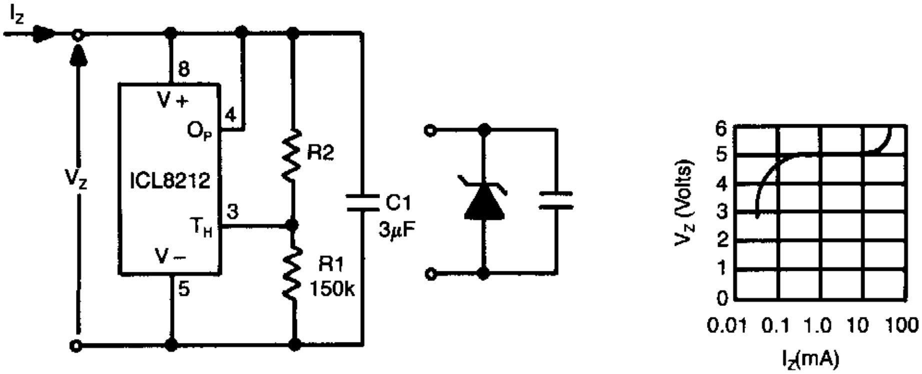
Fig. 74-1(b)

The special characteristic of this regulator is that the output voltage can be adjusted down to 0 V. The regulation is provided by an integrated regulator Type LM317. As is normal in supplies that can be adjusted to 0 V, this IC is used in conjunction with a zener diode. This diode provides a reference voltage that is equal, but of opposite sign, to the reference voltage (U_r) of the regulator, as shown in Fig. 74-1(a). Potential divider R1/R2 enables the output voltage to be adjusted.

In this circuit, the negative reference voltage is derived in a different manner: from the regulator with the aid of an op amp (Fig. 74-1(b)). The op amp is connected as a differential amplifier that measures the voltage across R1 and inverts this voltage to U_r. An additional advantage of this method is that at low-output voltages, a change in the reference voltage has less effect on the output voltage than the circuit in Fig. 74-1(a). The prototype, constructed as shown in Fig. 74-1(c), gave very satisfactory results.

The op amp need not meet any special requirements: a μA741 works fine, although an LF356 gives a slightly better performance. The negative supply for the op amp can be obtained with the aid of a center-tapped mains transformer.

PROGRAMMABLE ZENER



INTERSIL

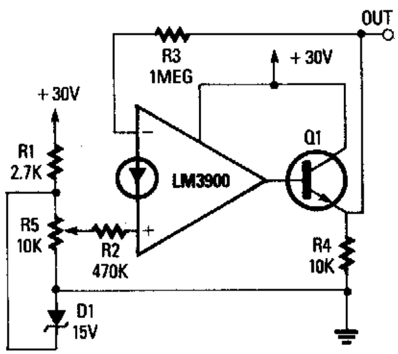
Fig. 74-2

The ICL8212 is connected as a programmable zener diode. Zener voltages from 2 V up to 30 V can be programmed by suitably selecting R2. The zener voltage is:

$$V_Z = 1.15 \times \frac{R_1 + R_2}{R_1}$$

Because of the absence of internal compensation in the ICL8212, C1 is necessary to ensure stability. Two points worthy of note are the extremely low-knee current (less than 300 μA) and the low dynamic impedance (typically 4 to 7 ohms) over the operating current range of 300 μA to 12 mA.

VARIABLE VOLTAGE REGULATOR



RADIO-ELECTRONICS

Fig. 74-3

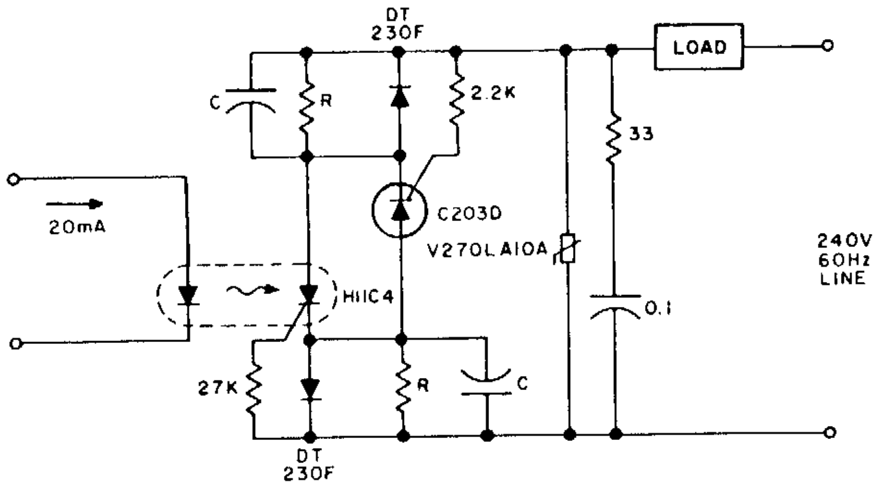
The op amp is wired as a ×2 noninverting dc amplifier with a gain that is determined by the R3/R2 ratio. The input voltage to the op amp is variable between 0 and 15 V via R5. The output voltage is therefore variable over the approximate range from 0.5 to 30 V. The available output current has been boosted by adding transistor Q1 to the output.

75

Relay Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 676. The figure number in the box of each circuit correlates to the entry in the Sources section.

LATCHING ac SOLID-STATE RELAY

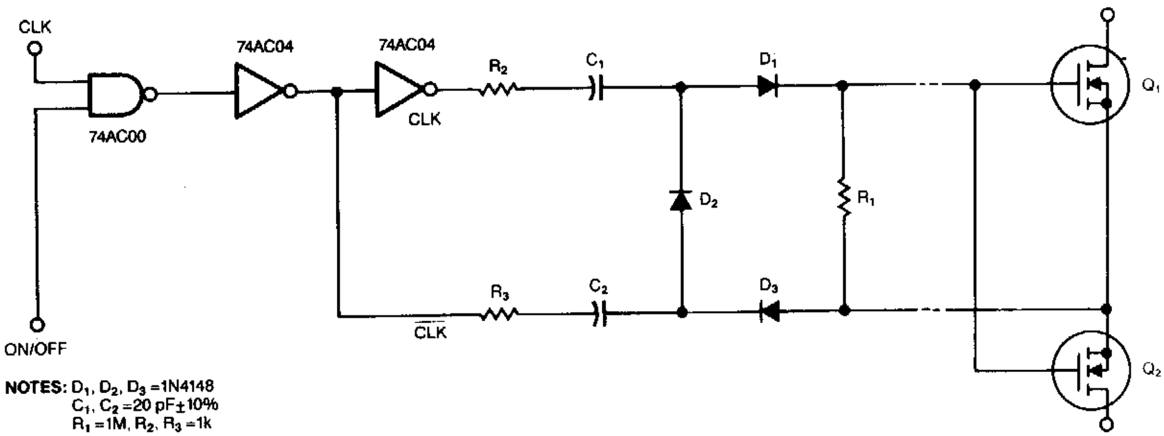


GE

Fig. 75-1

Latching is obtained by storing the gate trigger energy from the preceding half cycle in the capacitors. Power must be interrupted for more than one full cycle of the line to ensure turn-off. Resistor R and capacitor C are chosen to minimize dissipation, while assuring triggering of the respective SCRs for each cycle. A pulse of current, over 10 ms duration into the H11C4 IRED, ensures triggering the latching relay into conduction.

BIDIRECTIONAL SWITCH



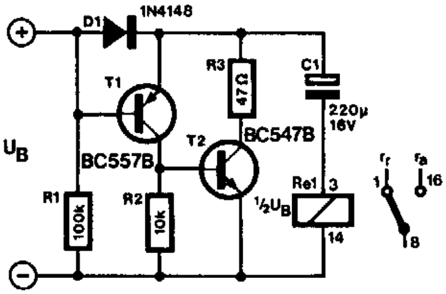
NOTES: D_1, D_2, D_3 = 1N4148
C_1, C_2 = 20 pF ± 10%
R_1 = 1M, R_2, R_3 = 1k

EDN

Fig. 75-2

Using voltage doublers, this simple switch circuit uses a clock signal and a control signal to switch MOSFETs.

LOW-CONSUMPTION MONOSTABLE RELAY



ELEKTOR ELECTRONICS

Fig. 75-3

A monostable relay has two states: *operative* when a large enough current flows through its coil and *quiescent* when no current flows. A relay contact that assumes a certain position after the supply voltage has been switched on is required in many applications. Of course, many relays operate in that manner.

However, most of these relays require an energizing current of 50 mA or more and that normally precludes a battery supply. The circuit presented here, which uses a bistable relay, can solve that problem.

The contact of a bistable relay normally remains in the position it is in after the supply is switched off. This circuit, however, makes the bistable relay behave like a monostable type, at a modest current.

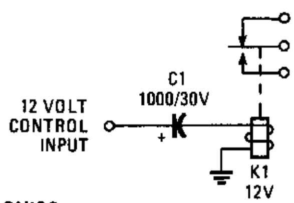
When the supply voltage is switched on, C1 charges via D1 and the relay coil. The current then flowing through the coil causes the relay contact to assume one of two positions. The forward drop across D1 ensures that the base of T1 (in this condition) is more positive than its emitter so that T1, and thus T2, is switched off.

When the supply voltage is switched off, the emitter of T1 is connected to the positive terminal of C1, while the base is connected to the negative terminal of the capacitor via R1 and the relay coil. This results in T1, and thus T2, switching on so that C1 discharges via T4 and the relay. The current flows through the relay coil, then flows in an opposite direction and this causes the contact to change over.

The bistable relay thus behaves exactly as a monostable with the advantage, however, that the operational current is determined by R1, which amounts to only 130 μA.

To ensure reliable operation, the rating of the relay coil should be 65 to 75% of the supply voltage. In the prototype, a 9-V relay was used with a battery supply voltage of 12 V.

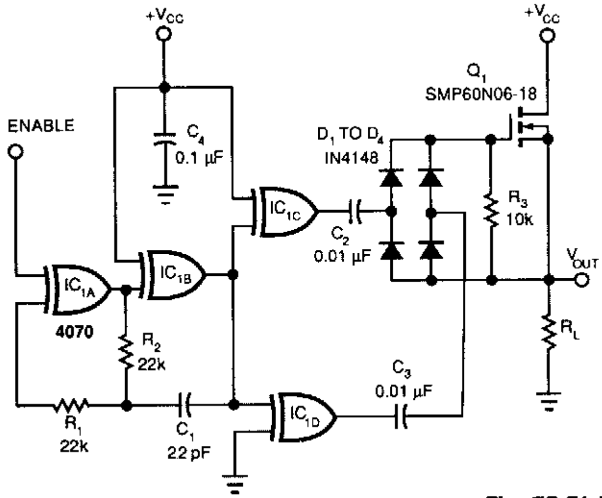
DELAY-OFF RELAY CIRCUIT



POPULAR ELECTRONICS

Fig. 75-4

When voltage is applied to the capacitor, it charges. While it's charging, the relay remains latched. When the charging current falls below the level needed to hold the relay down, the relay unlatches. The higher the value of the capacitor, the longer the relay will remain latched.



EDN

Fig. 75-5(a)

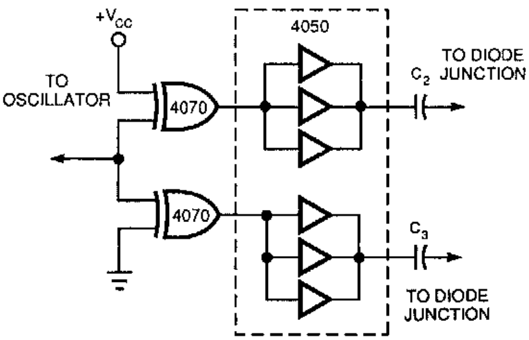


Fig. 75-5(b)

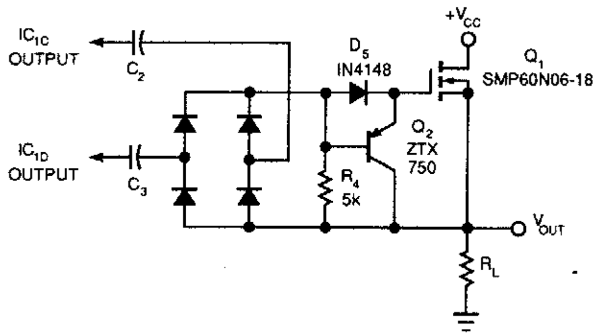


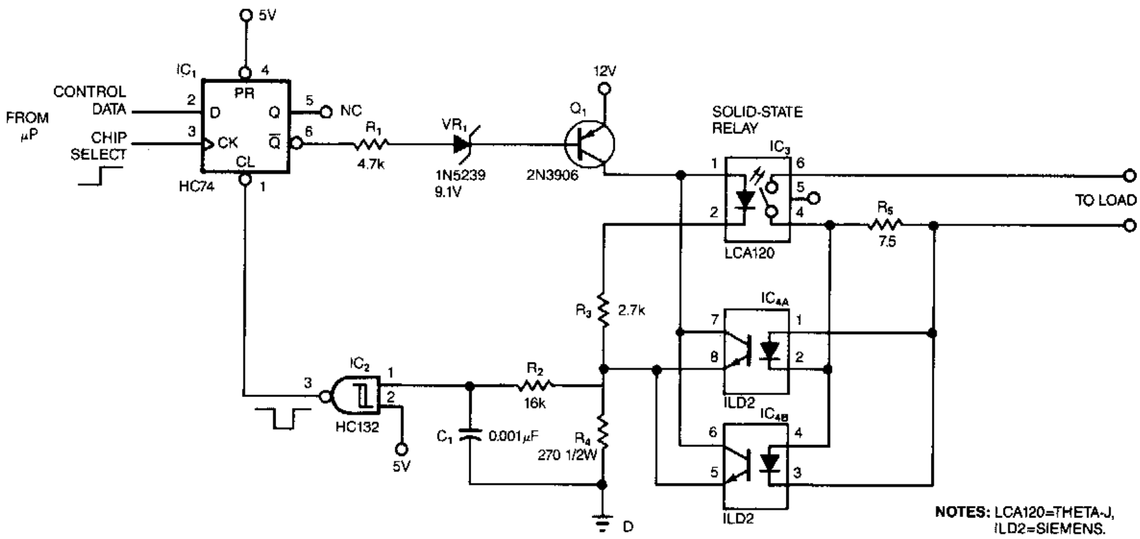
Fig. 75-5(c)

A power MOSFET and a quad exclusive-OR oscillator makes an effective solid-state relay. Figure 75-5(a)'s capacitively isolated drive circuit provides gate drive to turn on the n-channel device. This consists of a gated oscillator (IC1A and IC1B running at 500 kHz, set by R1, R2, and C1).

The diode bridge (D1 through D4) rectifies the charge transferred through C2 and C3. When you disable the oscillator, R3 discharges the stored gate charge, thereby turning off the MOSFET. R3 needs to allow fast turn-off times without loading the gate's enhancement voltage. A value of 10-kΩ is sufficient to produce a turn-off time of 800 μs with an 18-mΩ SMP60N06-18 MOSFET. The measured turn-on time is 150 μs.

You can reduce the turn-off time to 100 μs by using a pnp transistor as a diode-steering emitter-follower in the MOSFET gate circuit (Fig. 75-5(b)). Adding a hex buffer to Fig. 75-5A's circuit increases the drive capability of the complementary outputs (Fig. 75-5(c)).

OPTOISOLATOR



EDN

Fig. 75-6

The circuit protects a solid-state relay from overloads. The circuit limits current, automatically disconnects the load after detecting a short circuit, and develops a fault-condition output signal.

In normal operation, the controlling μP sets the flip-flop, IC1, which turns on transistor Q1. When Q1 turns on, current flows through the solid-state relay's input , thus activating the relay.

If an overcurrent or fault condition occurs, the excessive load current flowing through the relay develops enough potential across sense resistor R5 to turn on one of the optoisolators, IC4A or IC4B. The optoisolator's output transistor diverts current around the solid-state relay's input, which limits the current that the relay's output can pass.

If the overload is severe enough, the optoisolator pulls the input of the Schmitt trigger above its threshold, thus clearing the flip-flop and turning off the solid-state relay. R2 has two functions: It keeps the input of the Schmitt trigger below 5 V max. to prevent latchup, and it forms an RC filter in conjunction with C1. The RC filter prevents spurious triggering of the Schmitt trigger. You can use the output of the flip-flop to signal overload conditions to the controlling μP.

76

RF Amplifiers

The sources of the following circuits are contained in the Sources section, which begins on page 676. The figure number in the box of each circuit correlates to the entry in the Sources section.

4-W AMPLIFIER FOR 900 MHz

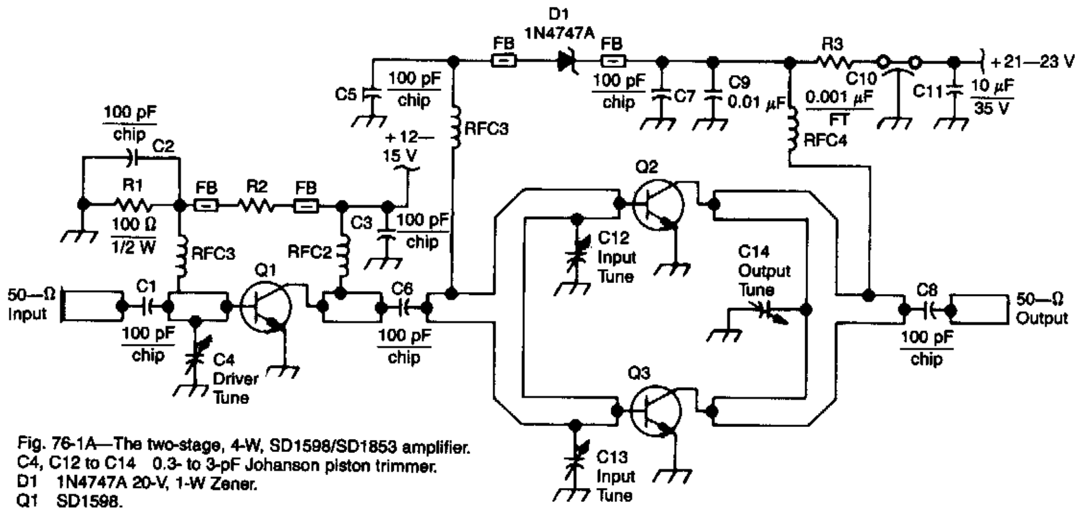
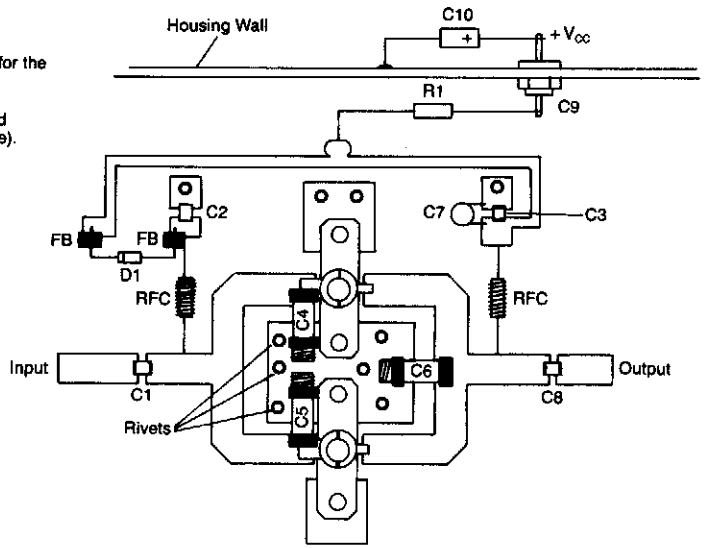


Fig. 76-1A—The two-stage, 4-W, SD1598/SD1853 amplifier.
C4, C12 to C14 0.3- to 3-pF Johanson piston trimmer.
D1 1N4747A 20-V, 1-W Zener.
Q1 SD1598.
Q2, Q3 SD1853.
R2 1.5 to 1.7 kΩ, 1/2 W. See text, Part 1, June 1990 *QST*, p 24.
R3 1 to 2 Ω, 2 W.
RFC1 to RFC4 8 turns of #26 enam. wire, closewound, 0.1″ ID.

Fig. 76-1B—Parts-placement diagram for the 2 × SD1853 amplifier. The PC-board edges are not shown. All components mount to the trace side of the PC board (except those mounted to the enclosure).

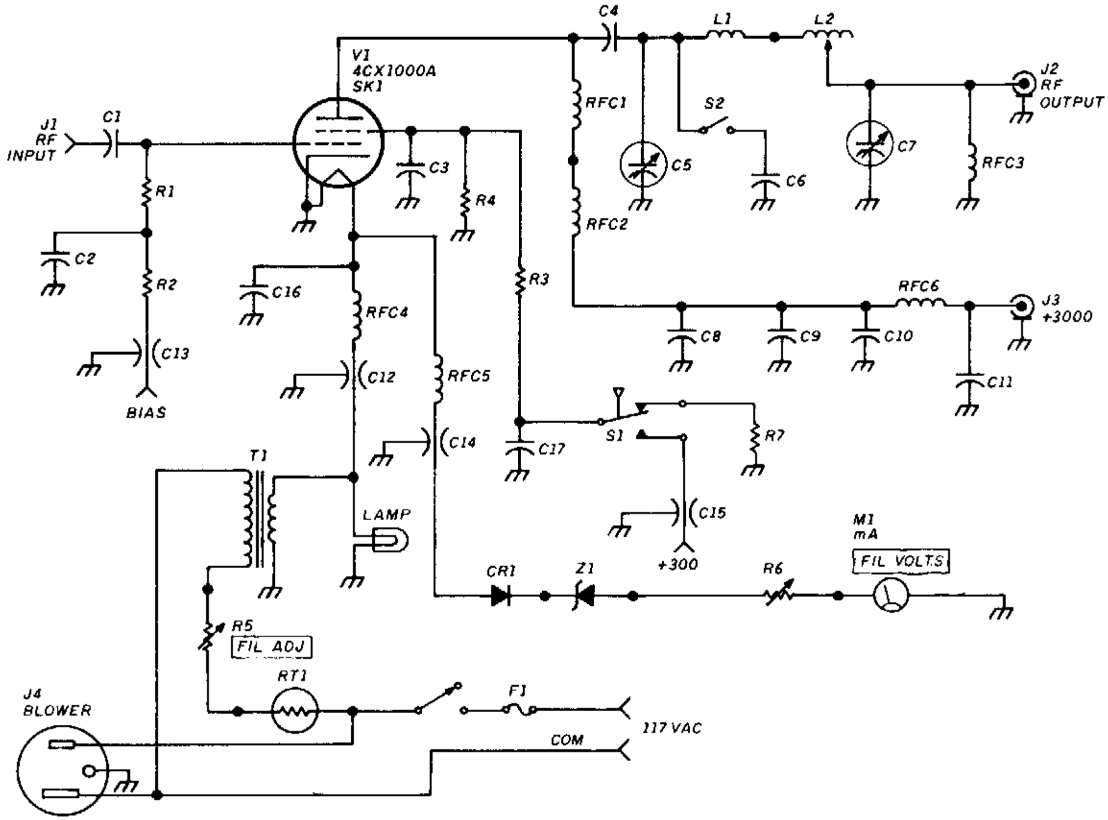


QST

Fig. 76-1

Using Wilkinson power dividers in the base and collector circuits of Q2 and Q3, two SD1853 devices are paralleled for twice the power output of the 2-W amplifier.

1500-W RF AMPLIFIER



C1,2—0.01 μF silver mica transmitting capacitor
C3—1500 pF screen grid bypass (internal to tube socket)
C4—2000 pF, 5 kV, two 858S capacitors in parallel, Centralab
C5—Vacuum variable, 10 to 300 pF, 10 kV, Jennings UCS-300 (plate tuning)
C6—300 pF, 5 kV, Centralab-type 858S
C7—Vacuum variable, 3000 pF maximum, 3 kV, Jennings UCSL-3000 (loading)
C8—0.0014 μF, 10 kV (EM)
C9—0.005 μF, 15 kV oil capacitor
C10,11—500 pF, 20 kV "TV doorknob," Centralab
C12—0.05 μF, 20 A feedthrough, Sprague
C13,14,15—1000 pF feedthrough
C16—0.1 μF, 100 volts DC disc ceramic
D1—Silicon diode, 1N4148 or 1N914
F1—3 A fuse
J1—BNC chassis mount
J2—N chassis mount
J3—HN chassis mount, used with RG213 for 3 kV feed
J4—Chassis-mount AC outlet, Amphenol 160-2N, AL part no. 713-5202
Lamp—No. 47 6.3 volt lamp
L1—3.5 turns no. 10 silver-plated wire, 1 inch ID, 2 inches long, self-supporting
L2—24 H roller inductor, Johnson 226-1 (EM, CC)
M1—1 mA DC meter movement, filament voltage
R1—50 ohm, 60 watt noninductive resistor, thirty 1500 ohm 2 watt carbon composition resistors connected in parallel (see text)
R2—1000 ohm, 2 watt carbon composition
R3—100 ohm, 2 watt carbon composition
R4—220 K, 2 watt carbon composition
R5—25 ohm, 25 watt wirewound adjustable
R6—1000 ohm, 2 watt carbon or wirewound pot

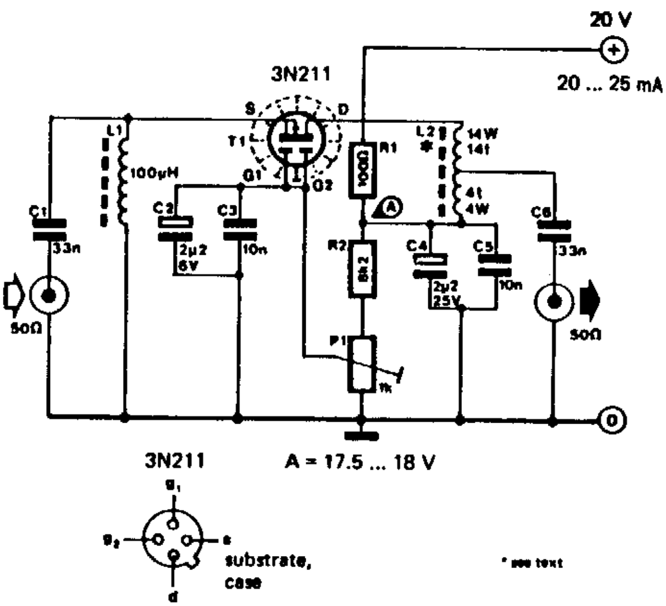
R7—10,000 ohm, 2 watt carbon composition
RFC1—28 turns no. 18 wire solenoid wound on 0.5 inch OD by 2.5 inch ceramic form (plate choke)
RFC2—Plate choke, surplus part (suggest B&W 801)
RFC3—74 turns no. 20 wire solenoid wound on 0.75 OD by 3 inch ceramic form
RFC4—9 μH 15 A RF choke, surplus part (suggest Dale no. IH15, HF no. 18-105)
RFC5—1.0 mH RF choke
RFC6—10 H, 1 A RF choke
RT1—Surgistor, GC 25-933-S
S1—Air flow switch, Rotron 2A-1350
SK1—Eimac socket, Sk-810B; chimney SK-806 (BY)
T1—Transformer, 6.3 volts AC at 10 A sec-ond—ary, Thordarson 21F12, AL 704-2019
Z1—Zener diode, 6.2 volts, 1N473

CQ

Fig. 76-2

The frequency range of this amplifier is 1.8 to 54 MHz. The amount of RF drive required for full output is about 30 W. The grid compartment (R1, R2, RFC4, RFC5) should be shielded from the other circuitry—especially the output circuitry.

WIDEBAND RF AMPLIFIER

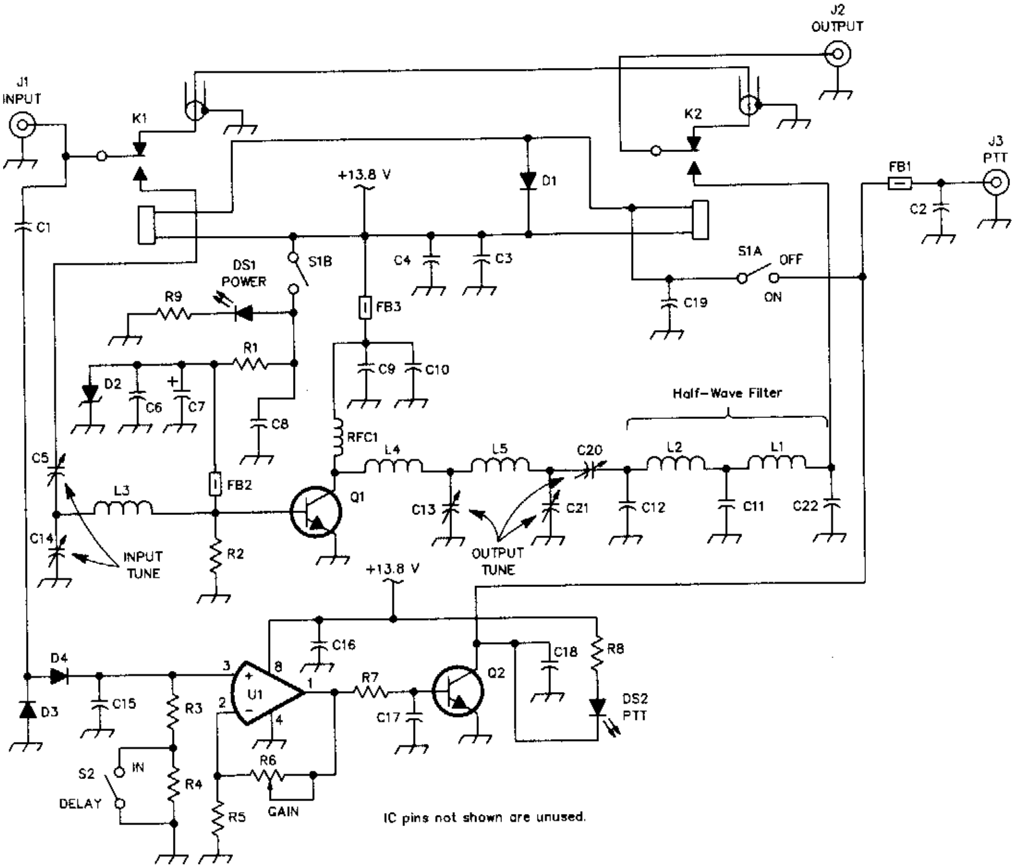


ELEKTOR ELECTRONICS

Fig. 76-3

This RF amplifier has wide bandwidth and dynamic range. It has a gain of 10 dB, noise figure less than 5 dB, and the third-order 1 MD is −40 dB at +22 dBm/tone output. The bandwidth is 4 to 55 MHz.

6-m 100-W LINEAR AMPLIFIER



C1—10 pF, ceramic disc.
C2—0.01 μF, polyester film.
C3, C10—0.001 μF, ceramic disc.
C4, C6, C8, C9, C15-C19—0.1 μF, polyester film.
C5—9- to 180-pF mica compression trimmer, Arco no. 463.
C7—1000 μF, 6.3 V, aluminum electrolytic.
C11—120 pF, 100 V, silver mica.
C12, C22—62 pF, 100 V, silver mica.
C13, C14, C20—50- to 380-pF mica compression trimmer, Arco no. 465.
C21—25- to 280-pF mica compression trimmer, Arco no. 464.
D1—1N4002.
D2—1N1200 stud-mount diode.
D3, D4—1N4148.
DS1—Green LED.
DS2—Red LED.

FB1, FB3—Ferrite bead, Amidon no. FB43-901 or similar.
FB2—VK-200 wide-band choke, 2½ turns no. 24 solid wire on Amidon no. FB43-5111 ferrite core.

J1, J2—Female RF connector (UHF, BNC, N, etc).
J3—Phono jack.
K1, K2—SPDT relay, 12-V dc coil, Omron no. G5L112P-PS-DC12. Available from Digi-Key.
L1, L2—4 turns no. 14 enam wire, 7/16 in. diam, 3/8 in. long.
L3—2 turns no. 14 enam wire, 3/8 in. diam, ½ in. long.
L4—no. 14 U-shaped wire loop, 3/8 in. diam, 9/16 in. finished length.
Q1—MRF492.
Q2—2N2222A.
R1—5-W bias resistor (see text).
R2—10 Ω, ½ W, carbon comp.
R3, R5—10 kΩ, ¼ W, carbon comp.
R4—1 MΩ, ¼ W, carbon comp.
R6—100-kΩ PC-board potentiometer.
R7-R9—1 kΩ, ¼ W, carbon comp.
S1—DPDT miniature toggle.
S2—SPST miniature toggle.
U1—LM358.

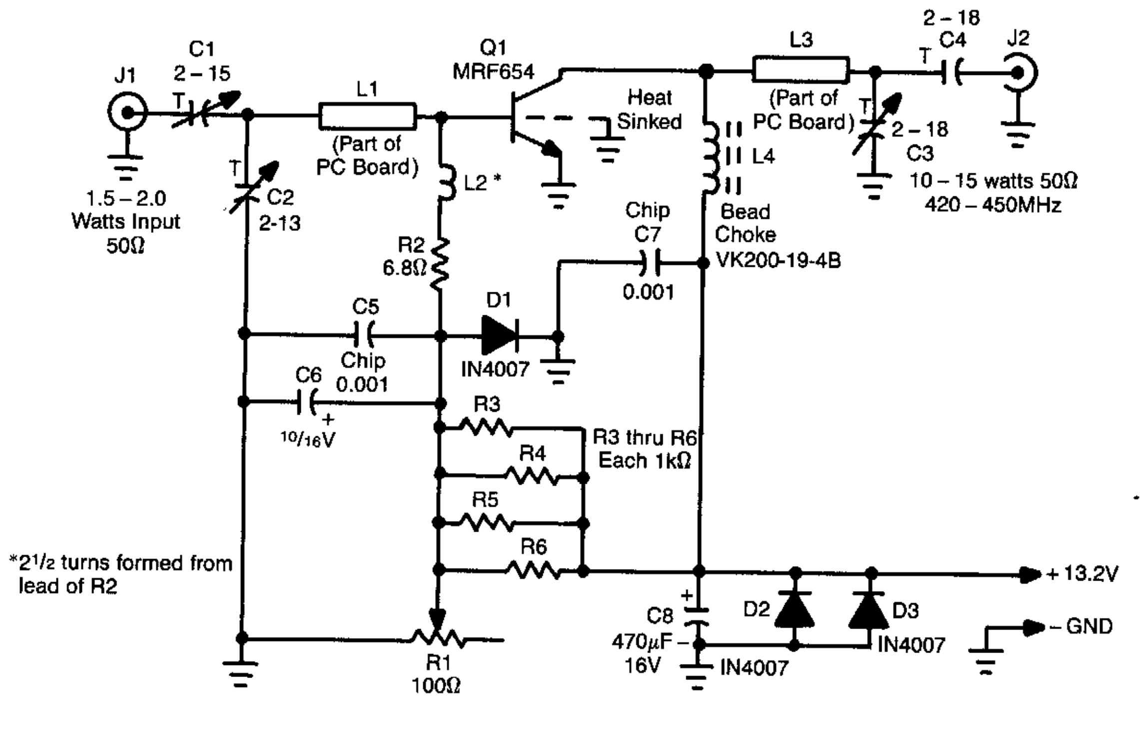
QST

Fig. 76-4

Miscellaneous
Suitable die-cast-aluminum enclosure and heat sink.
Pair of binding posts or other suitable dc-supply connector.
Two 1- × 3-in. strips of double-sided foam tape.
Two LED holders.
Seven no. 4-40 × ¼ in. machine screws.

100 W output at 50 MHz is available from this circuit. U1 and Q2 form a T-R relay driver, switching the amplifier on when RF input at J1 is sensed. During receive periods, J1 and J2 are directly connected. A 13.8-V supply is required for this amplifier.

10- to 15-W ATV LINEAR AMPLIFIER

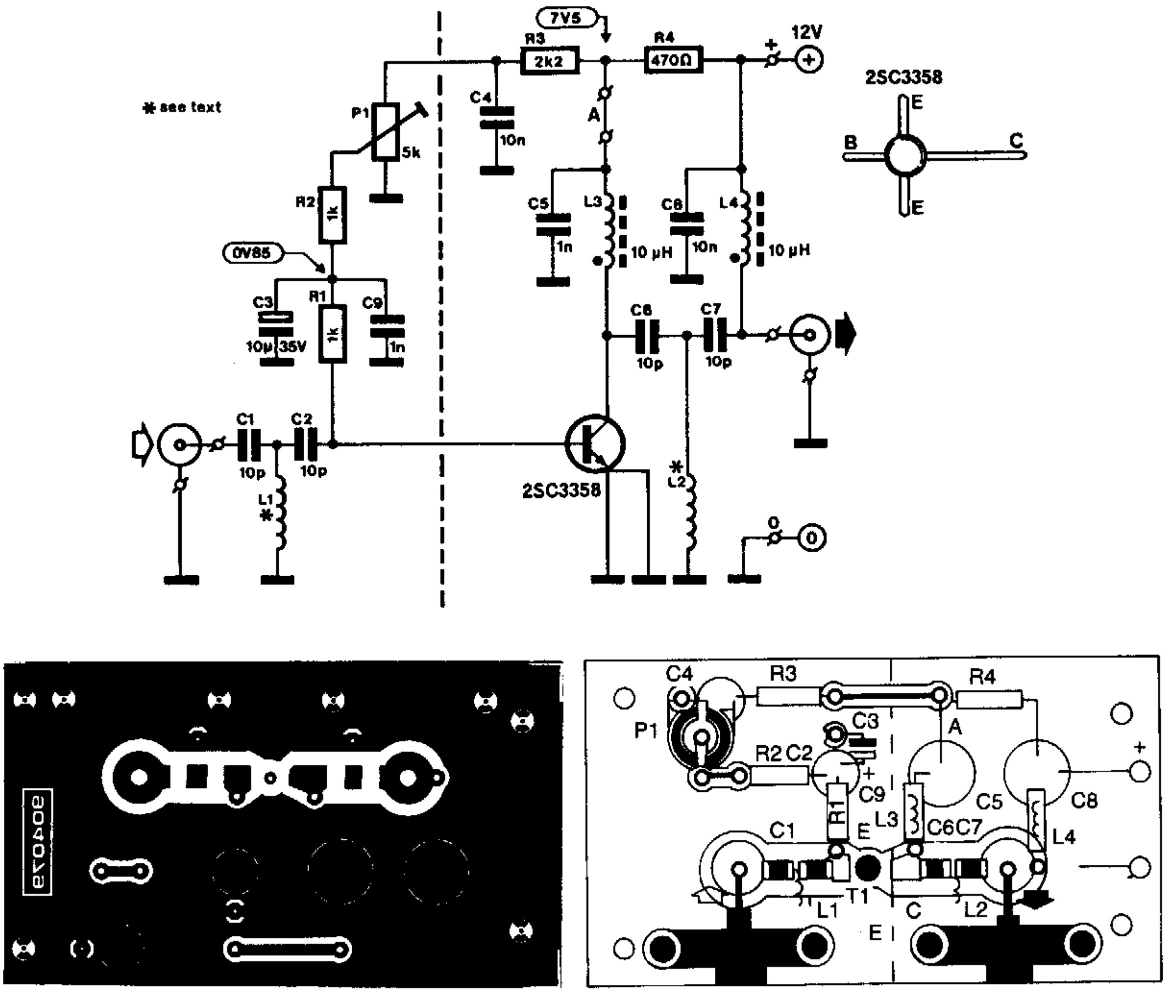


NORTH COUNTRY RADIO

Fig. 76-5

This amplifier is useful for applications where a 10- to 15-W peak-envelope-power (PEP) signal is needed in the 420- to 520-MHz range. C1, C2, and L1 form a matching network for amplifier Q1. L3, C3, and C4 form an output matching network for a 50-Ω load. L2, R2, D1, and R3 through R6 (with bias adjust R1) form a biasing network for Q1. D2 and D3 provide polarity protection for Q1, which must be heat-sinked. A kit of all parts including case and PC board is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804.

UHF TV-LINE AMPLIFIER



ELEKTOR ELECTRONICS

Fig. 76-6

This circuit offers 10- to 15-dB gain from 400 to 850 MHz and is therefore eminently suitable for situations where the television signal is weak. Moreover, the filters can be adapted to the individual needs of users. Construction is simplicity itself if the ready-made PC board shown on the next page is used. The tracks should be tinned or silvered for optimum performance and long life.

The opening at the center of the board is intended to accommodate the transistor. This device has two emitter pins, both of which should be connected to ground. The drawings show that the board is divided into two by a small piece of tin plate, which should have a small cut-out for the transistor.

The input and output terminals are made from small cable clamps and M3 nuts and bolts. One side of disc capacitors C4, C5, C8, and C9 is soldered direct to the board.

Input and output capacitors, C1/C2 and C6/C7 respectively are surface-mount types. C1/C2/L1 form an input filter and C6/C7/L2 form an output filter. The value of the capacitors might have to be lowered to

3.9 pF to obtain the correct frequency range. The amplifier can be housed in a watertight case and then mounted near the antenna at the top of the mast (if used).

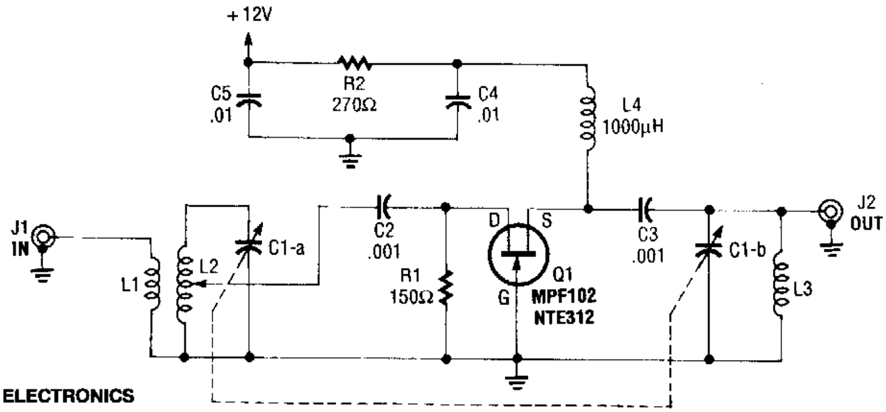
The power is obtained from a simple stabilized 12-V supply: a mains adapter with a 78L12 will do nicely. This can be kept indoors, of course. The amplifier can be powered via the coaxial feeder cable, for which purpose a 10- to 100-μH choke is inserted in the supply line. Calibrating the amplifier is straightforward: set P1 to the center of its travel and then adjust it for optimum picture quality.

In practice, the collector current of the transistor is then 5 to 15 mA. This may be checked by temporarily replacing jump lead A by one suitable meter.

PARTS LIST

R1, R2 = 1 kΩ
R3 = 2 kΩ
R4 = 470 Ω
P1 = 5 kΩ preset potentiometer
C1, C2, C6, C7 = 10 pF surface mount
C3 = 10 μF; 35 V
C4, C8 = 1 nF disc
C5, C9 = 1 nF disc
L1, L2 = air core, 2 turns of 3mm diameter enamelled copper wire
L3, L4 = 10-μH choke or 10 turns of 0.2 mm diameter enamelled copper wire on a ferrite bead.
T1 = 2SC3358

DOUBLE-TUNED JFET PRESELECTOR



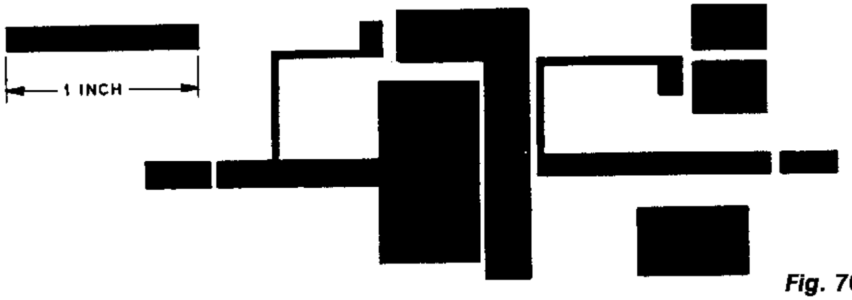
POPULAR ELECTRONICS

Fig. 76-7

This circuit uses an MPF102 JFET and a double-tuned common-gate amplifier. Gain is typically 10 to 15 dB:

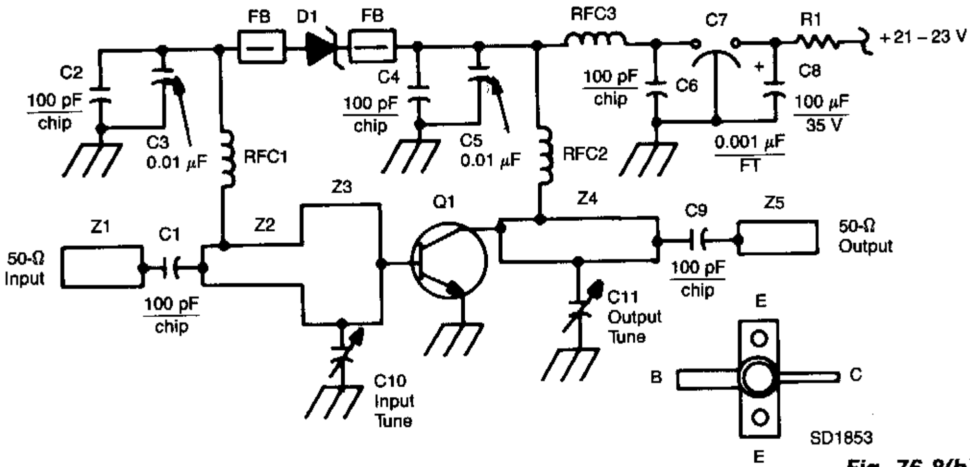
$$L_2 \text{ and } L_3 = \frac{1}{(2\pi f)^2 C_1}$$

L1 ≈10% turns on L2 or 1 turn (larger of these two).



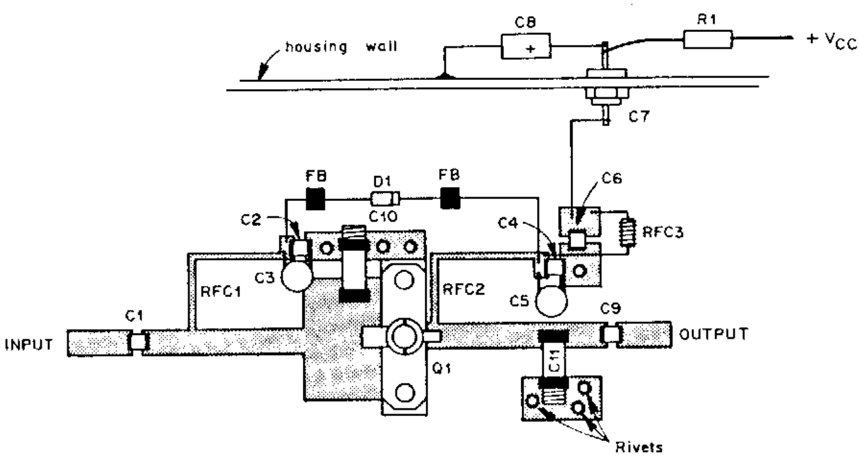
QST

Fig. 76-8(a)



QST

Fig. 76-8(b)



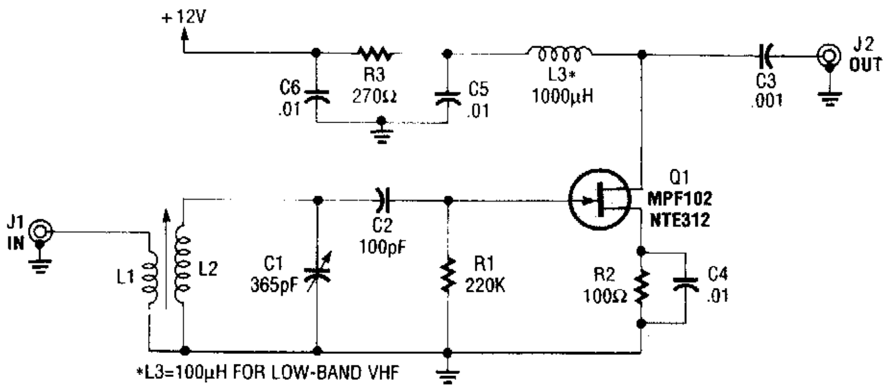
QST

Fig. 76-8(c)

The 2-W SD1853 amplifier. RFC1 and RFC2 are implemented as PC-board traces.
C10,C11 0.3- to 3-pF Johanson piston trimmer.
D1 1N4747A 20-V, 1-W Zener.
Q1 SD1853.
R1 2 – 3 Ω, 1 W.
RFC3 8 turns of #26 enam wire, closewound, 0.1″ ID.
Z1 – Z5 Microstriplines. See text and Fig. 18.

Two W output can be produced by this circuit designed for the 902- to 928-MHz amateur band. As shown in the figures, much of the circuitry is in the form of PC board traces.

SIMPLE JFET PRESELECTOR



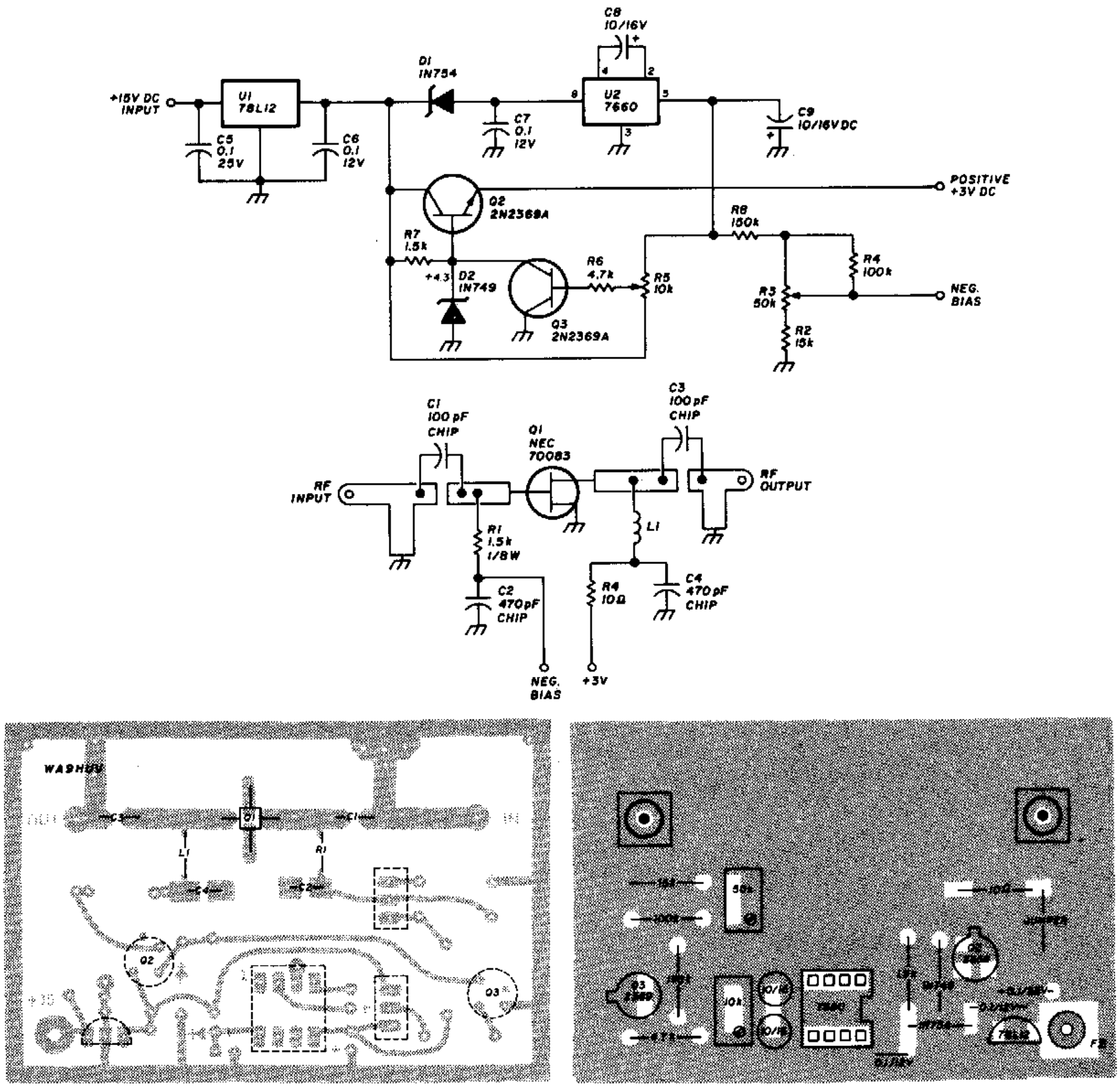
POPULAR ELECTRONICS *Fig. 76-9*

This preselector will improve the performance of simple shortwave receivers. L2 is tuned with Q1. The inductance of L2 is:

$$L_2 = \frac{1}{(2\pi f)^2 C_1}$$

and L1 is found around the cold end of L2. Typically, L1 has 10% of the turns in L2, or one turn, whichever is larger.

1296-MHz RF AMPLIFIER



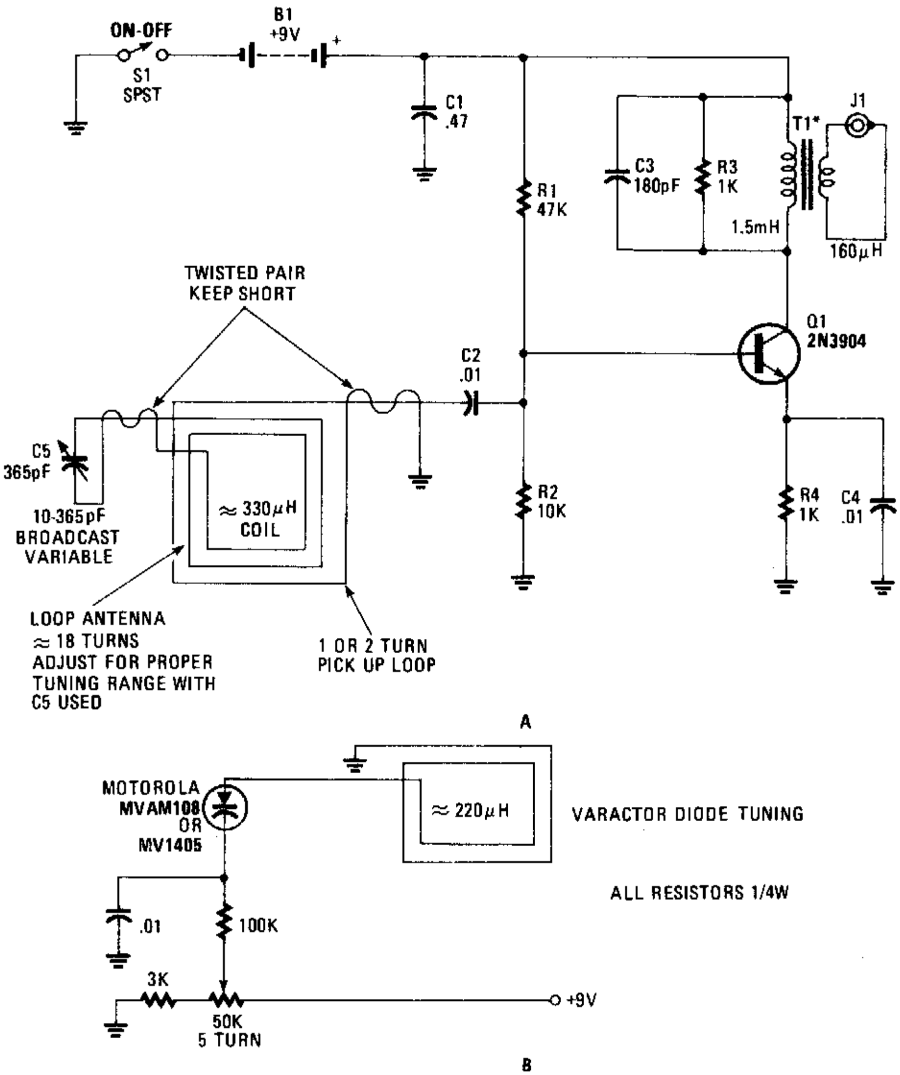
HAM RADIO

Fig. 76-10

Using an NEC70083, this 1296-MHz amplifier delivers about 17-dB gain and around 1- to 1.5-dB noise. It is constructed on a G-10 epoxy fiberglass PC board. Use the layout shown because this is important for correct performance.

The power supply/regulator delivers the regulated 3-Vdc for the drain circuit and U2 produces a negative bias for the gate circuit. R5 sets the drain voltage to +3 V and R3 sets the gate bias. Typically, the drain current is about 10 mA.

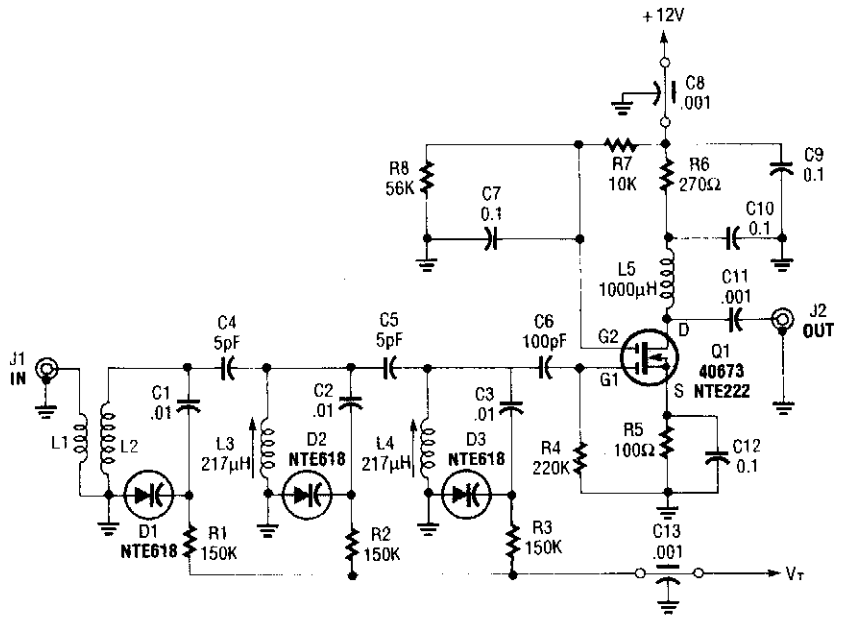
BROADCAST BAND BOOSTER



POPULAR ELECTRONICS *Fig. 76-11*

The use of a loop antenna of large size (12 × 18″) provides a large signal pickup at AM broadcast frequencies. It has about 18 turns. T1 is a toroidal transformer of about 3:1 turns ratio (not critical). The primary winding should have about 1.5 mH inductance. By using the circuit at (B), a varactor diode can be used in place of the 10- to 365-pF variable capacitor.

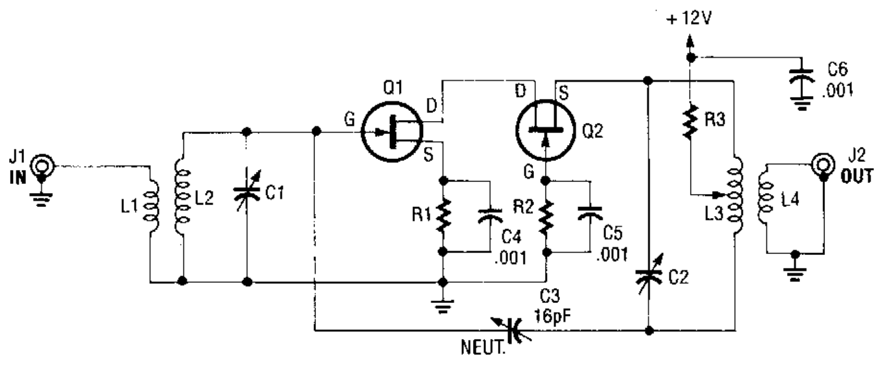
VARACTOR-TUNED PRESELECTOR



POPULAR ELECTRONICS *Fig. 76-12*

Varactor diodes replace the conventional 365-pF tuning capacitors, which reduces size and weight when used as a turned RF stage for AM broadcast applications. Selectivity is good enough for use as a TRF receiver if a detector is connected to J2.

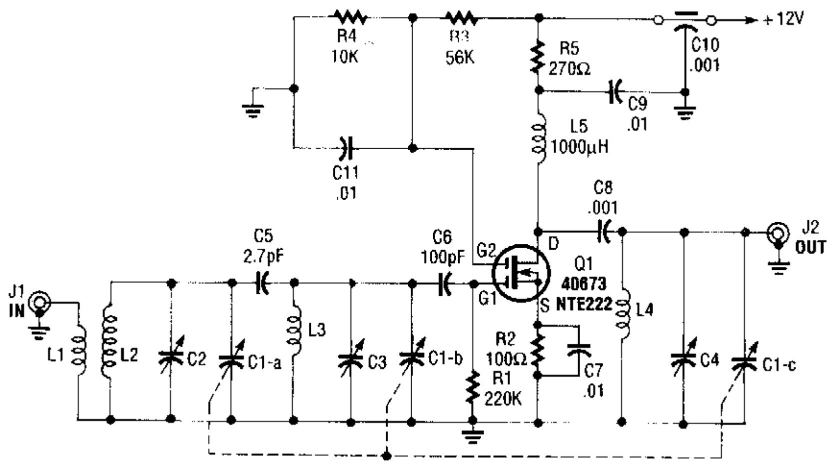
CASCODE RF AMPLIFIER



POPULAR ELECTRONICS *Fig. 76-13*

A cascode amplifier using two MOSFETs is shown in the diagram. L2C1 and L3C2 resonate to the frequency in use. The circuit has the advantage of good gain, low NF, and excellent linearity. Q1 and Q2 can be MPF102 or 2N4416.

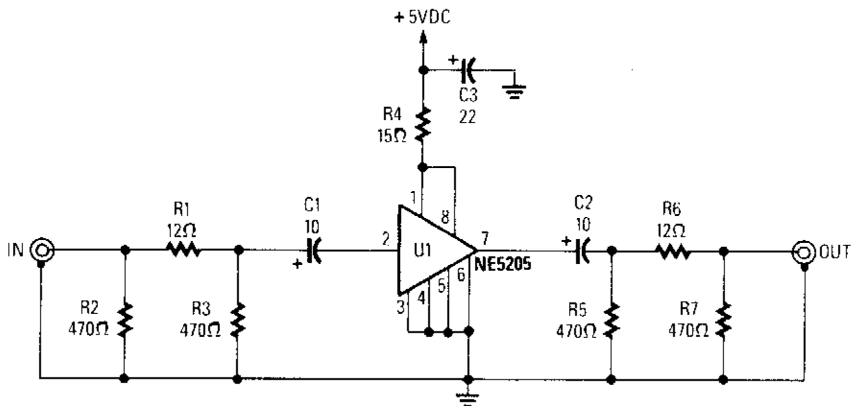
DUAL-GATE MOSFET RF-AMP STAGE



POPULAR ELECTRONICS *Fig. 76-14*

The use of a double-tuned input and a single-tuned output yield superior RF selectivity to that of equivalent single-tuned designs. AGC, if required, can be added to gate 2 of Q1, and should drive gate 2 negative for decreased gain.

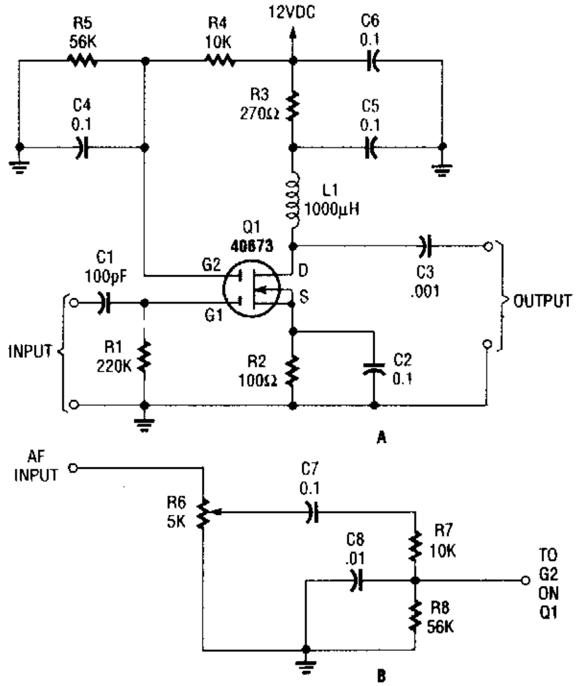
WIDEBAND AMPLIFIER



POPULAR ELECTRONICS *Fig. 76-15*

Using a Signetics NE5205, this circuit gives about +16-dB gain from LF to 600 MHz. The π minimizes load impedance and source-impedance variations and aids stability.

BUFFER AMPLIFIER WITH MODULATOR

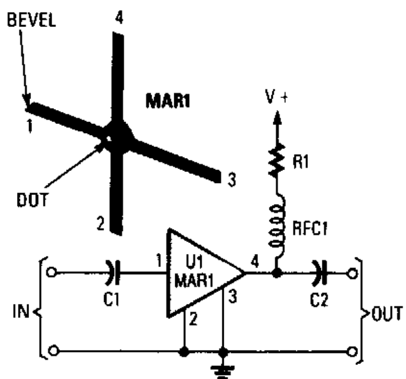


POPULAR ELECTRONICS

Fig. 76-16

A 40673 MOSFET is used as a wideband buffer amplifier (to 20 MHz). If desired, the amplifier can be modulated, considering that the gate 2 voltage of a MOSFET can be used to vary the gain of the stage.

WIDEBAND AMPLIFIER

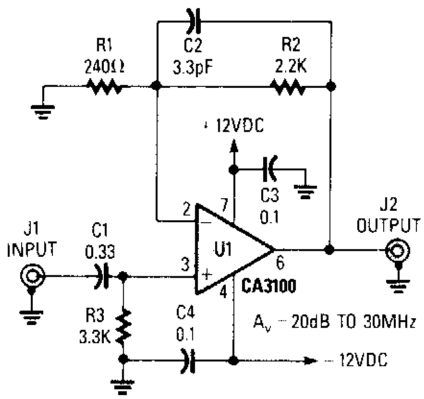


POPULAR ELECTRONICS

Fig. 76-17

This wideband amplifier uses a MAR1 IC, which is a gain block. The device is manufactured by Mini-Circuits Lab and offers +13-dB gain from dc to 1 000 MHz. R1 is selected to provide 17-mA current to U1 at +5 V. For 12-V supply $R_1 = 470\ \Omega$. RFC1 is typically 1 to 5 μH.

TWO CA3100 WIDEBAND OPERATIONAL AMPLIFIERS



POPULAR ELECTRONICS *Fig. 76-18(a)*

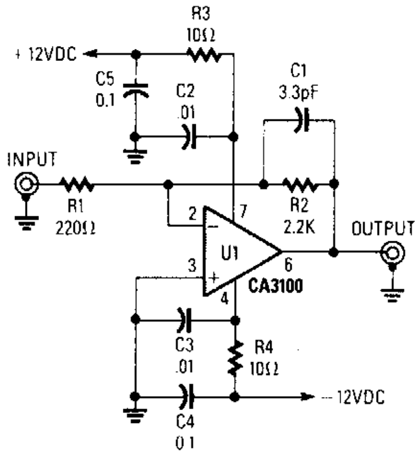


Fig. 76-18(b)

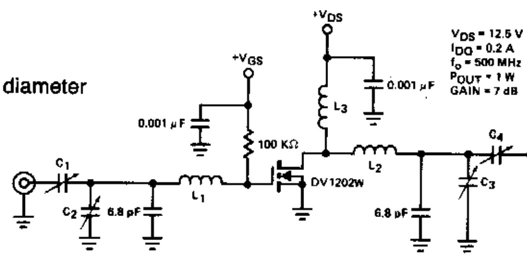
These circuits use the RCA CA3100 wideband amplifier IC. The gain is:

$$A_v = 1 + \frac{R_2}{R_1}$$

These circuits are useful for video applications to 10 MHz. The 3.3-pF capacitors are compensation capacitors. Capacitors on pins 7 and 4 are bypass capacitors to prevent self-oscillations. Figure 76-18(a) is noninverting and Fig. 76-18(b) is an inverting configuration.

500-MHz AMPLIFIER

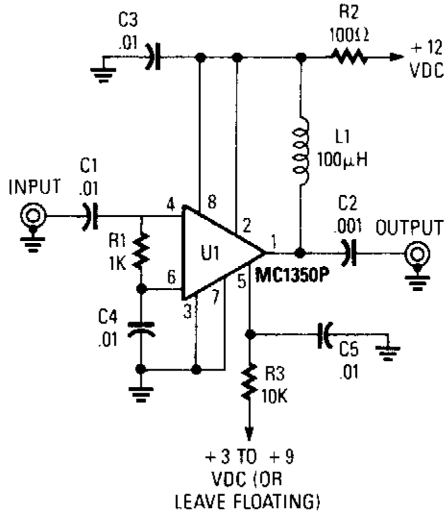
C1, C2, C3, C4, ARCO #400, 1→7pF
L1, L2, 1/2″ length #12 wire
L3, 4 turns #22 enameled wire close wound on 1/4″ diameter



MOTOROLA *Fig. 76-19*

This amplifier provides 1-W output at 500 MHz with a DV1202W FET. About 6- to 8-dB power gain can be expected.

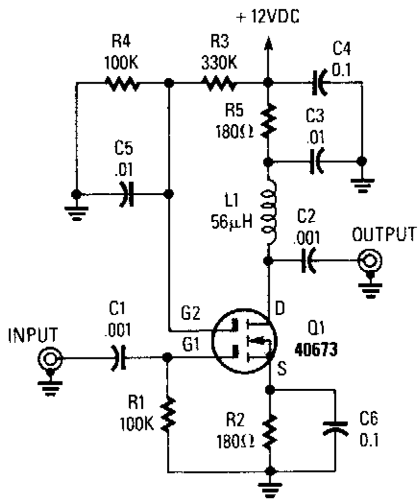
HF WIDEBAND AMPLIFIER



POPULAR ELECTRONICS *Fig. 76-20*

About 20-dB gain can be obtained using this IC. The gain can be controlled by varying the voltage applied to R3.

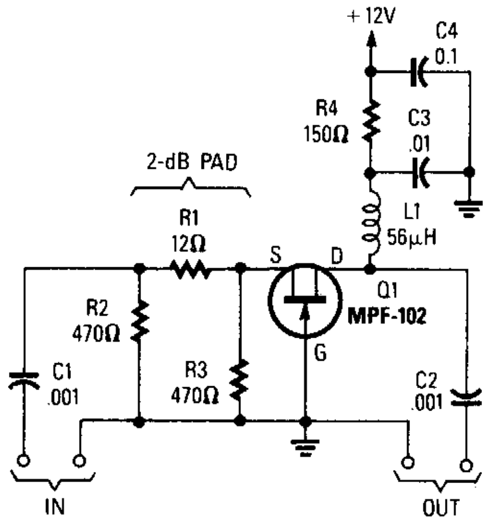
MOSFET WIDEBAND AMPLIFIER



POPULAR ELECTRONICS *Fig. 76-21*

For high-impedance (7 500 Ω) applications, this amplifier will provide a voltage gain of approximately $-gm/Z_L$ where Z_L is the load impedance in ohms and gm is $\approx 12 \times 10^{-3}$ for the 40673 FET. The G2 voltage can be used to control the gain.

JFET WIDEBAND AMPLIFIER



POPULAR ELECTRONICS *Fig. 76-22*

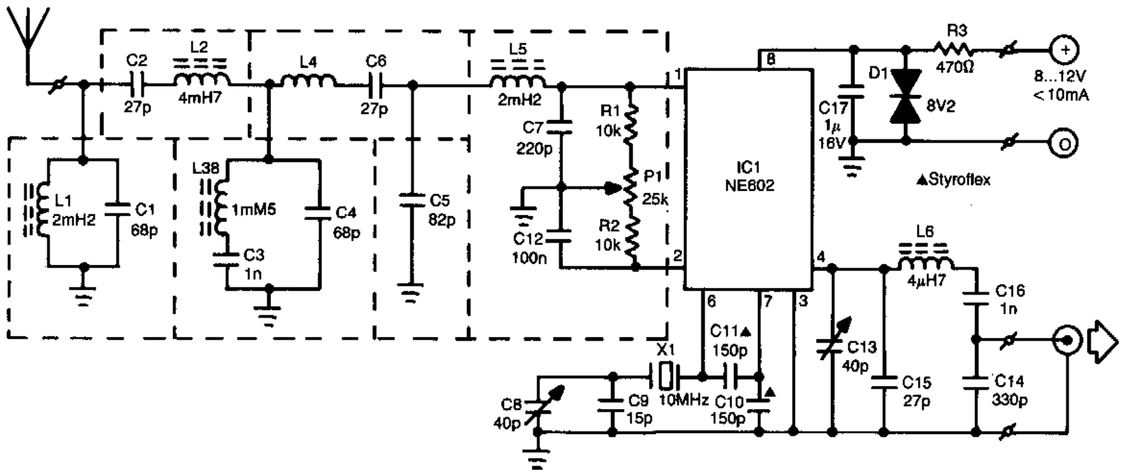
Using an MPF102 JFET, this circuit has a 50-Ω input impedance. Load impedance should be about 470 Ω. A 3:1 matching transformer can be used to get the impedance back to 50 Ω. This circuit will show about 6- to 8-dB gain at VHF frequencies.

77

RF Converters

The sources of the following circuits are contained in the Sources section, which begins on page 677. The figure number in the box of each circuit correlates to the entry in the Sources section.

RADIO BEACON CONVERTER



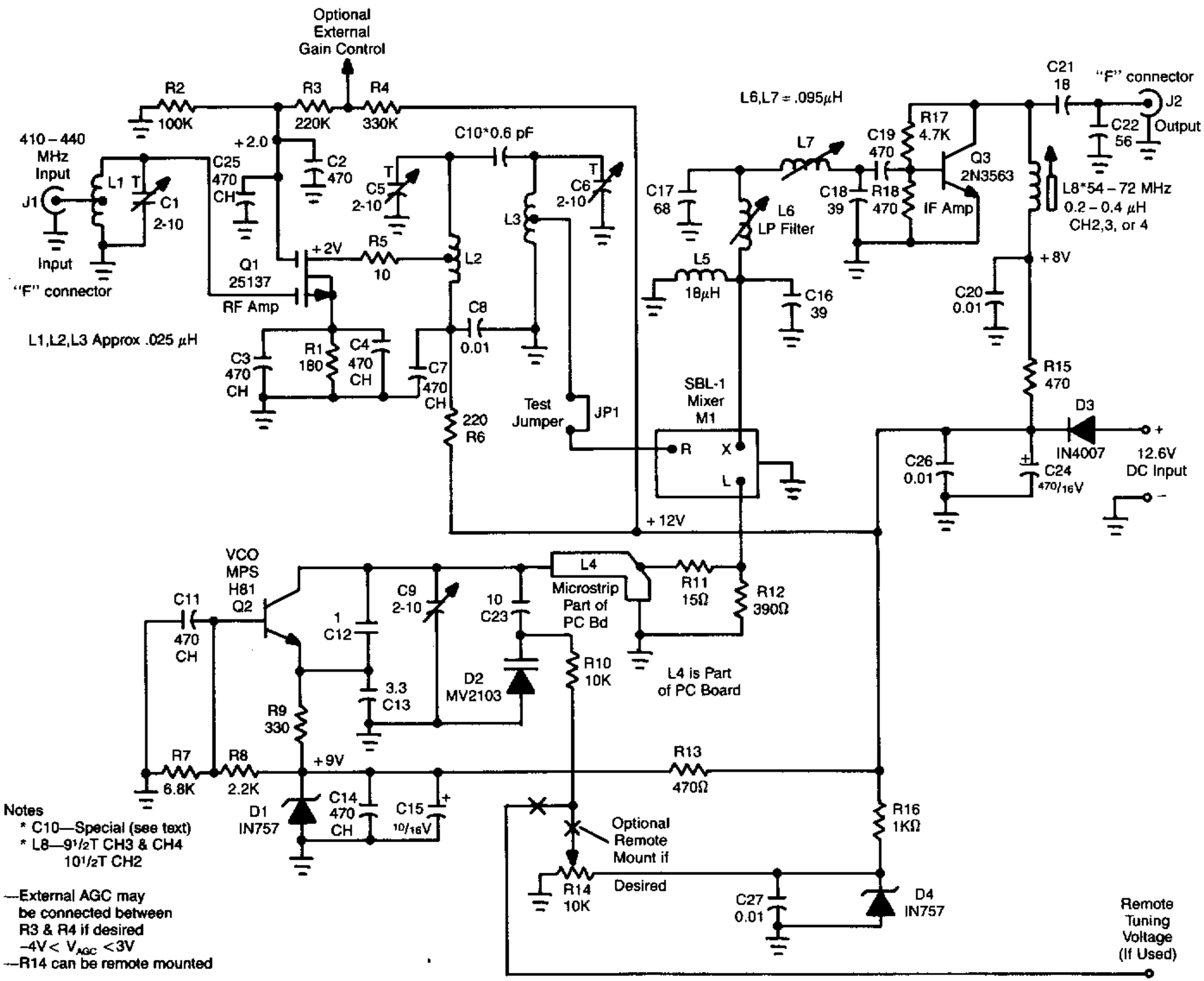
ELEKTOR ELECTRONICS *Fig. 77-1*

The radio beacon band extends from 280 to 516 kHz. Each beacon has its own characteristic AM-modulated morse-coded callsign that is transmitted on a specific frequency. To be able to receive distant beacons, the aerial signal is passed through a band-pass filter that effectively suppresses longwave and mediumwave signals. The filter also converts the aerial impedance, Z_{in}, from about 10 kΩ to the input impedance of mixer IC1, which is about 1 kΩ.

The mixer adds or subtracts the received signal to/from the local oscillator signal so that the beacon signal can be received on a normal shortwave receiver. The resulting frequencies are from 9.72 to 9.48 MHz or from 10.280 to 10.516 MHz. In the construction of the converter, some components must be surrounded by a metal shield, as indicated by dashed lines on the PC board layout.

The circuit is aligned with the aid of an SSB receiver, to which the output of the converter is connected. Tune the receiver to 10 MHz and adjust the oscillator frequency of the converter with C8 for zero beat. Next, detune the receiver slightly until you hear a pleasant whistle, which is adjusted for minimum level with the aid of P1. Finally, tune to a beacon transmitting at or about 300 kHz and adjust C13 for maximum sound output.

LOW-NOISE 420-MHz ATV RECEIVER/CONVERTER



Notes
* C10—Special (see text)
* L8—9½T CH3 & CH4 10½T CH2

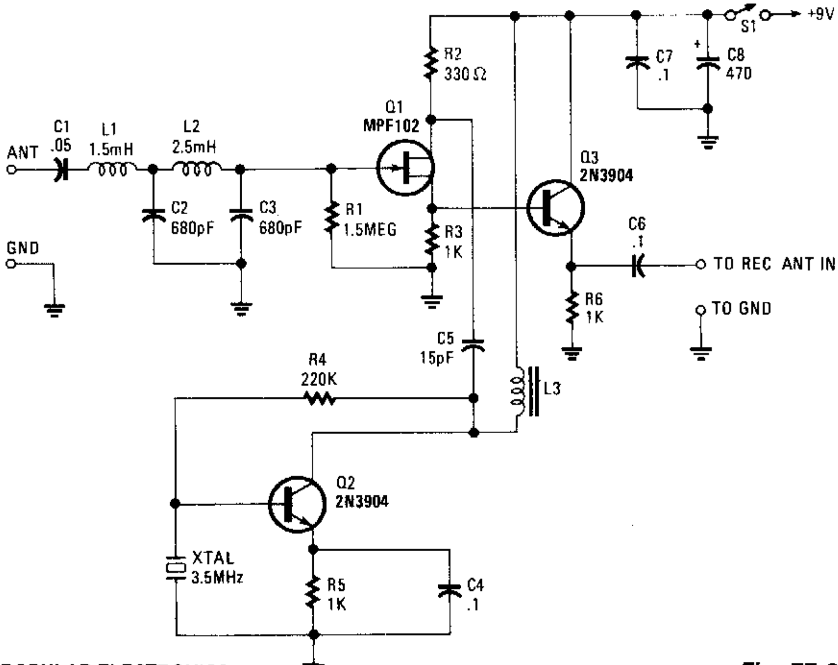
—External AGC may be connected between R3 & R4 if desired $-4V < V_{AGC} < 3V$
—R14 can be remote mounted

NORTH COUNTRY RADIO

Fig. 77-2

L1, Q1, L2, and L3 compose an RF amplifier stage that feeds M1, a doubly balanced mixer. Q4 is a local oscillator stage in the 375-MHz range. Signals in the 420- to 450-MHz range from Q1 are mixed in M1 and fed through filter L6/L7/C17, where only the 60- to 70-MHz (CH3/CH4) signals pass. The IF signal is passed to Q3, an IF amplifier. The overall gain is 25 dB and the noise figure less than 2 dB. A kit of all parts, including the PC board, is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804.

VLF CONVERTER

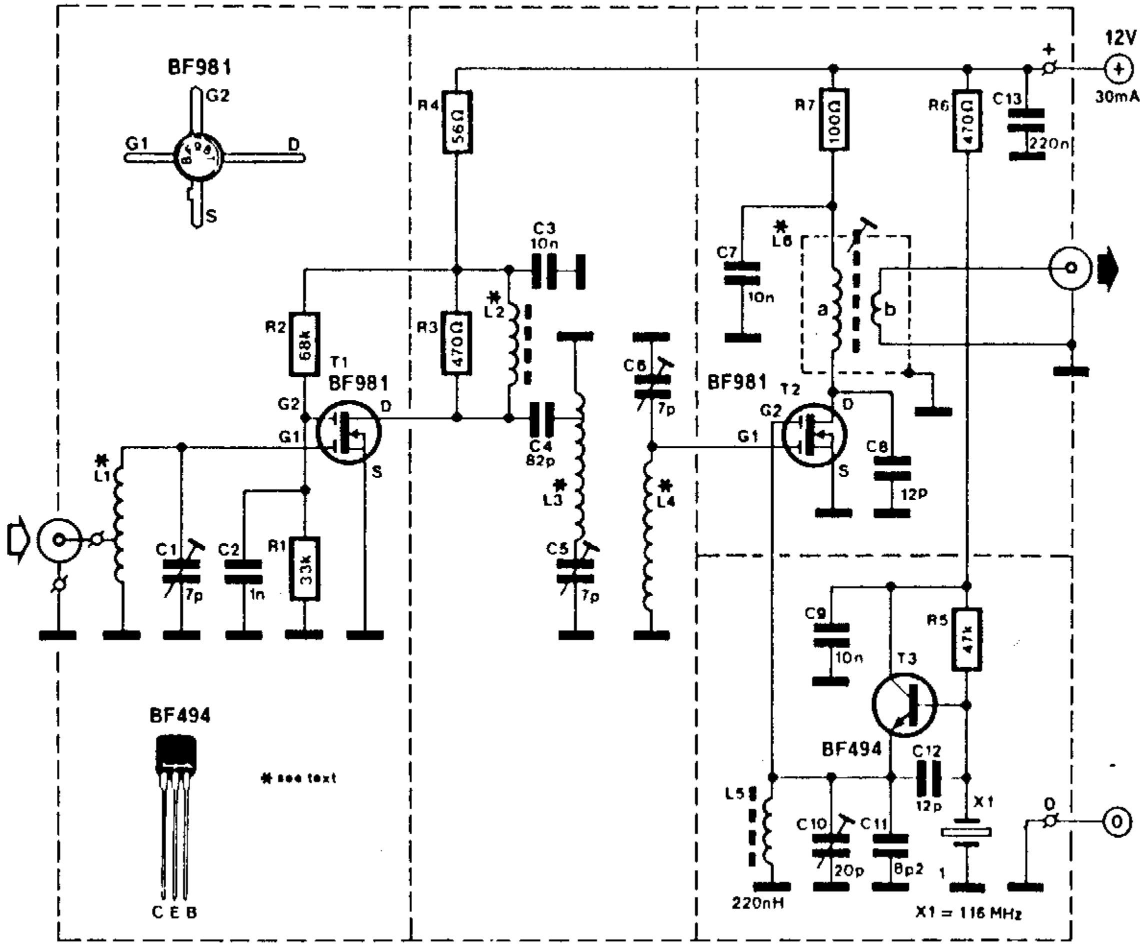


POPULAR ELECTRONICS *Fig. 77-3*

The VLF Converter can be used to pick up signals for the general coverage of shortwave receivers. A number of unusual signals can be heard on frequencies below 15 kHz.

This converter will convert frequencies from 0 to 250 kHz to 3 500 to 3 750 kHz so that the LF- and VLF-band segments can be received on an amateur or shortwave receiver that covers 3 500 to 4 000 kHz. Signals from a short whip antenna (8 to 10 feet) are coupled through low-pass filter L1/L2/C2/C3 to RF amp Q1. Q3 mixes these signals with a 3.5-MHz signal from Q2 and associated components C4, R5, R4, and 3.5-MHz XTAL. L3 is an RF choke that presents an inductive load to Q3. It should be resonant somewhat above 3.5 MHz when placed in the circuit. An adjustable coil of about 30 to 100 μH should be sufficient. The converter output is taken from the emitter of Q3 through C6.

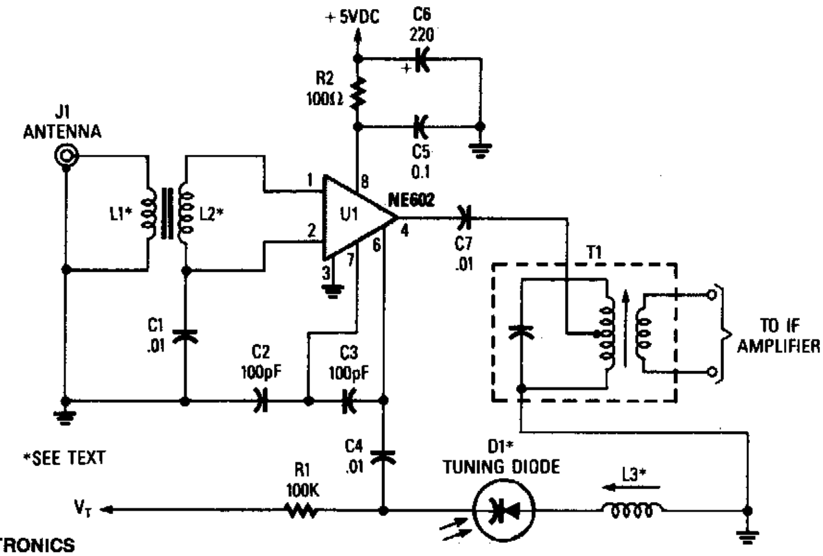
2-METER CONVERTER



ELEKTOR ELECTRONICS *Fig. 77-4*

This converter enables a receiver that tunes 28 to 32 MHz to receive the 144- to 148-MHz amateur band. A BF981 dual-gate MOSFET provides RF gain and feeds mixer T2, another BF981. T3 is a 116-MHz crystal oscillator used to provide L.O. injection to T2. Coils are wound on a 6-mm form. L1, L3, and L4 are 8 turns of 1-mm diameter silver-plated copper wire. L2 is 4 turns of 0.2-mm wire through a ferrite lead. L6 has 19 turns on the primary and 3 turns on the secondary.

RECEIVER FREQUENCY-CONVERTER STAGE



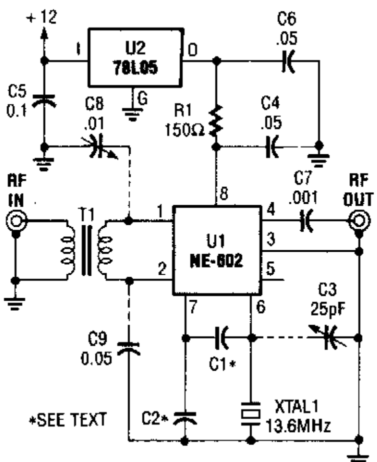
POPULAR ELECTRONICS

Fig. 77-5

L1, L2 1:12 Turns Ratio Toroid (Broadband).
L3 Resonates to L.O. Frequency with D1 capacitance.
LO FREQ Desired received frequency ± IF frequency.

In this case, the NE602 is used in this superhet front-end configuration. U1 serves as a frequency converter. L1/L2 is a broadband toroidal transformer. A tuned transformer may be used instead. The supply voltage is +5 to +9 Vdc. T1 is tuned to the IF frequency. The typical IF frequency is 455 kHz. This circuit, depending on L1, L2, and L3, should be usable in the frequency range from audio to 30 MHz. The varactor tuning diode can be replaced with an air-variable capacitor, if desired.

10-MHz WWV TO 80-METER SW CONVERTER



POPULAR ELECTRONICS

Fig. 77-6

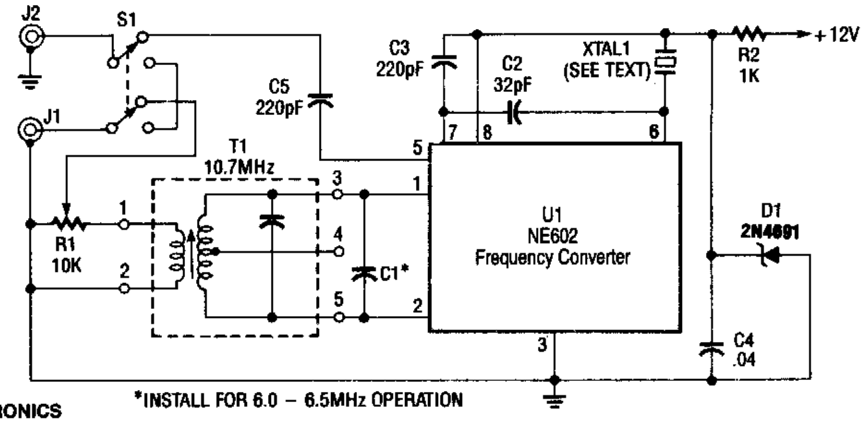
This converter is useful where reception of WWV is desirable and only a ham-band receiver is available. U1 acts as a mixer/oscillator. The values of C_1 and C_2 are given by:

$$C1 = \frac{100 \text{ pF}}{\sqrt{f_{MHz}}}$$

$$C2 = 10 \times C_1$$

$$f_{MHz} = \text{crystal frequency}$$

SHORTWAVE CONVERTER FOR AM CAR RADIOS

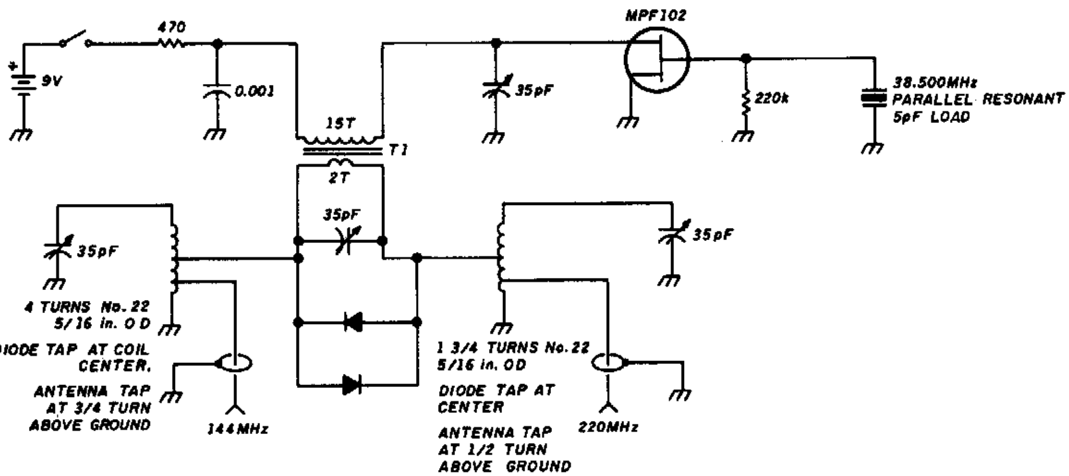


RADIO-ELECTRONICS

Fig. 77-7

Using a Signetics NE602, this converter tunes the 9.5- to 9.8-MHz range. An AM car radio is used as a tunable IF amplifier. Output is taken from J2, the auto antenna. The crystal (XTAL1) can be a frequency about 1 MHz below the desired tuning range; for 9.5 to 9.8 MHz, an 8.5- to 8.8-MHz crystal should be used.

220-MHz RECEIVING CONVERTER

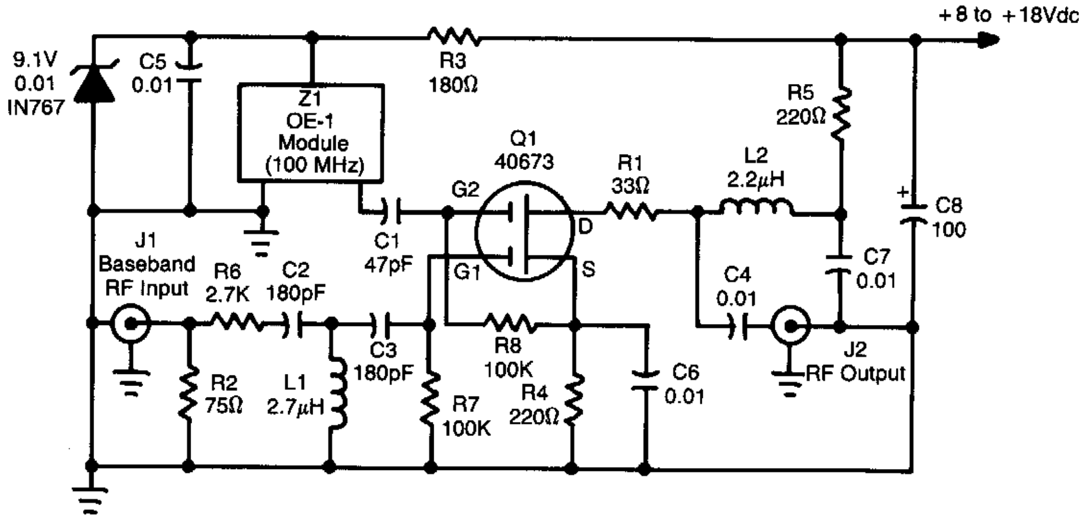


HAM RADIO

Fig. 77-8

A simple circuit using a single transistor converts 220 MHz to 144 MHz or vice versa, because the mixer is bilateral. T1 has 15 turns on the primary, and 2 turns on the secondary (#24 AWG wire) on a 0.375″ ID SF-material toroidal coil.

RF UPCONVERTER FOR TVRO SUBCARRIER RECEPTION



RADIO-ELECTRONICS

Fig. 77-9

This converter uses a 40673 MOSFET to heterodyne the 5.5- to 8-MHz TVRO subcarriers to the FM broadcast band, where a stereo receiver can be used for high-fidelity stereo reception of TV sound subcarriers. Z1 is a prepackaged 100-MHz oscillator module, which is available from International Crystal Corporation, Oklahoma City, OK.

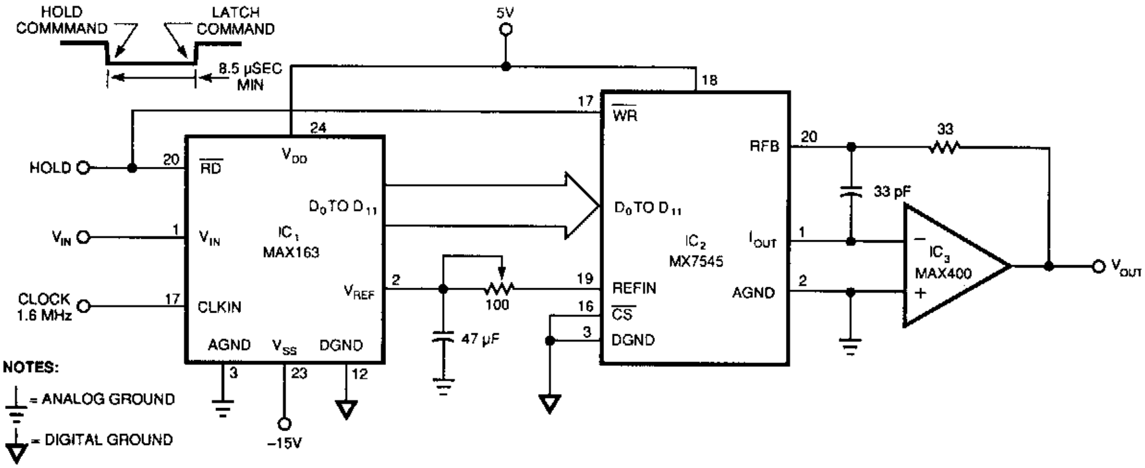
78

Sample-and-Hold Circuit

The source of the following circuit is contained in the Sources section, which begins on page 677. The figure number in the box correlates to the entry in the Sources section.

Sample and Hold

SAMPLE AND HOLD



EDN

Fig. 78-1

By using the ADC and DAC, this circuit uses a negative pulse to latch data into the ADC. The hold pulse width should be at least 8.5 μs. This circuit has zero drop and infinite hold time.

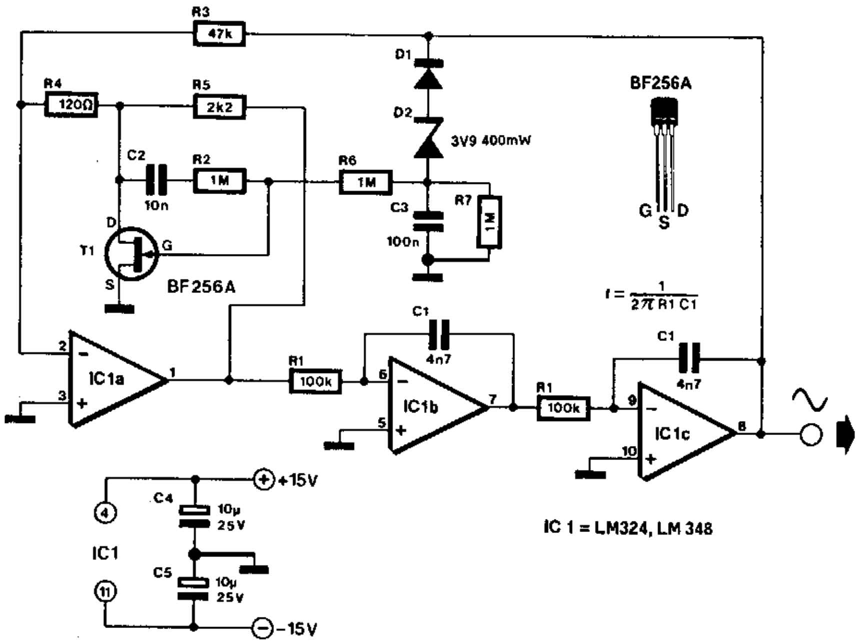
79

Sine-Wave Oscillators

The sources of the following circuits are contained in the Sources section, which begins on page 677. The figure number in the box of each circuit correlates to the entry in the Sources section.

Sine-Wave Generator
Pure Sine-Wave Generator
LC Sine-Wave Generator
60-Hz Sine-Wave Generator
Two-Transistor Sine-Wave Oscillator
VLF Audio Tone Generator
Very Low Distortion Oscillator
Low-Frequency LC Oscillator
Three-Decade 15-Hz to 15-kHz Wien-Bridge Oscillator
Phase-Shift Oscillator for Audio Range
Wien-Bridge Oscillator
Low-Frequency Sine-Wave Generator
Sine- and Square-Wave TTL Oscillator
Wien-Bridge-Based Oscillator with Very Low Distortion

SINE-WAVE GENERATOR



ELEKTOR ELECTRONICS

Fig. 79-1

The frequency of the generator is determined by integrators IC1B and IC1C. An integrator has two properties that are used in this design. Firstly, a phase shift of 90° is between the input and output (ignoring, for the moment, the nonideal behavior of the op amp), and secondly, its amplification is −1 (i.e., the signal inverts), provided the frequency:

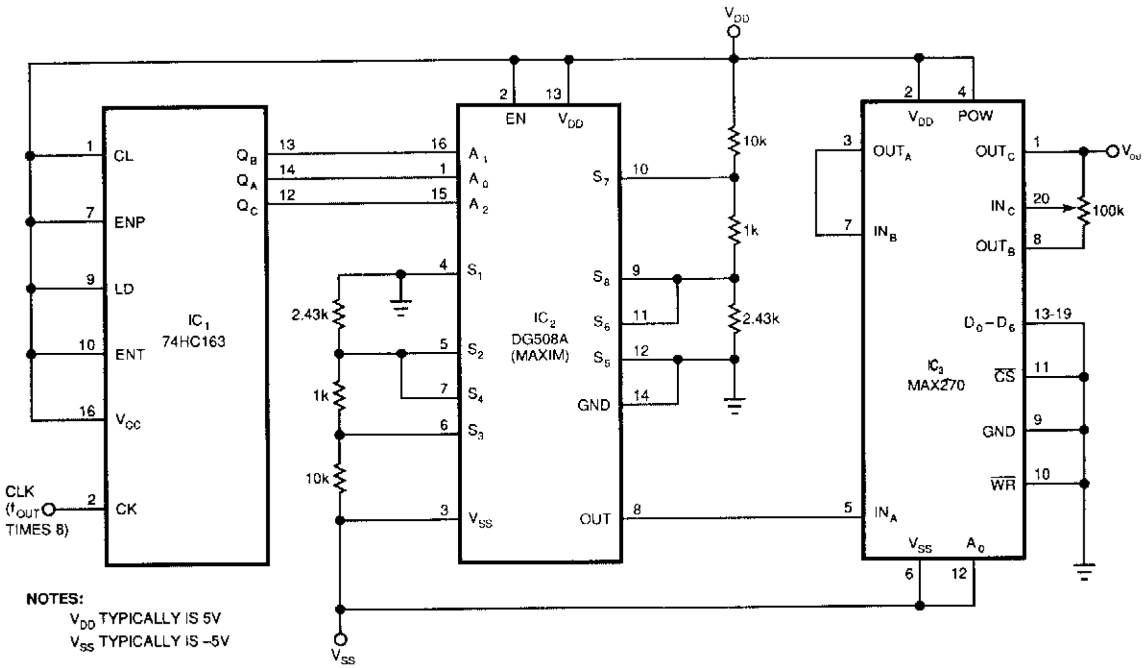
$$f = \frac{1}{2\pi R_1 C_1}$$

Cascading two identical integrators will thus result in an overall phase shift of 180° and an amplification of unity (provided that the frequency is $1/2\pi R_1 C_1$): an ideal basis for an oscillator. The two integrators are connected in the feedback circuit of an amplifier whose gain is determined by the amplitude of the output signal. Consequently, the generator has reasonably stable output voltage (at a level of about 4.5 Vpp).

With the values of C1 (C1′) and R1 (R1′), as shown in the diagram, the output has a frequency of about 300 Hz. The frequency can be varied by replacing R1 and R1′ with a stereo potentiometer. To keep the frequency setting within bounds, the overall range of this potentiometer should not exceed a decade.

The maximum attainable frequency is about 5 kHz. Distortion is not greater than 0.1%. The current drawn by the generator is only a few milliamperes. Finally, the LM348 is a quadruple 741; it is thus possible to construct the generator from four 741s.

PURE SINE-WAVE GENERATOR



EDN

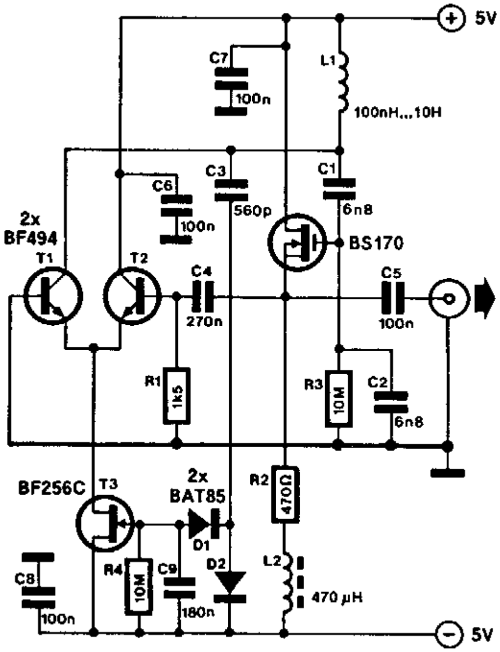
Fig. 79-2

This circuit produces a pure, −80-dB THD sine wave with a frequency that is equal to the f_c of IC3's filter. A TTL counter, an 8-channel analog multiplexer, and a fourth-order low-pass filter can generate 1- to 25-kHz sine waves with a THD of better than −80 dB. The circuit cascades the two second-order, continuous-time Sallen-Key filters within IC3 to implement the fourth-order low-pass filter. Two resistive dividers connected from ground to V_{DD} and ground to V_{SS} provide bipolar dc inputs to the multiplexer.

To operate the circuit, you first must choose the filter's cutoff frequency, f_c, by tying IC3's D_0 through D_6 inputs to 5 V or ground. The cutoff frequency can be at 128 possible levels between 1 and 25 kHz, depending on those 7 digital input levels. Because this figure ties D_0 through D_6 to ground, f_c equals 1 kHz. The 100-kHz potentiometer adjusts the output level anywhere from 1.5 V below V_{DD} to 1.5 V above V_{SS}.

The clock input frequency must be 8 times higher than the filter's f_c. The multiplexer then produces an 8× oversampled staircase approximation of a sine wave. 8× oversampling greatly simplifies the smoothing requirements of the low-pass filter by pushing the first significant harmonic out to 7× the fundamental. All higher-order harmonics are removed by IC3, which includes an uncommitted amplifier for setting the output level.

LC SINE-WAVE GENERATOR



ELEKTOR ELECTRONICS *Fig. 79-3*

This compact LC oscillator offers a frequency range of about 1 kHz to almost 9 MHz and a low-distortion sine-wave output. The heart of the circuit is series-resonant circuit L1/C2/C3 in the feedback loop of amplifiers T1/T2. Transistor T2, which is connected as an emitter follower, serves as impedance converter, whereas T1, connected to a common base circuit, is a voltage amplifier whose amplification is determined by the impedance of L1 in its collector circuit and the emitter current.

The feedback loop runs from the collector of T1 via the junction of capacitive divider C1/C2, source-follower BS170, and the input impedance is formed by R1/C4. The whole is strongly reminiscent of a Colpitts circuit. The signal is also taken to the output terminal via C5.

Of particular interest is the amplitude control by the current source. The signal is rectified by two Schottky diodes, smoothed by C9, then used to control the current through T3. The gain of amplifier T1 is therefore higher at low input levels than at higher ones. This arrangement ensures very low distortion, because the amplifier cannot be overdriven.

The resonant frequency can be calculated from:

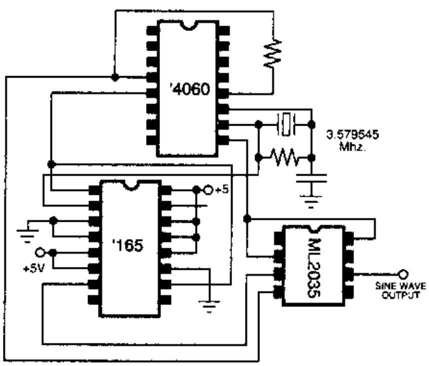
$$f = \frac{1}{2\pi}\left(\frac{L_1\,C_1\,C_2}{C_1 + C_2}\right)$$

With values as shown, it extends from 863 Hz ($L_1 = 10$ H) to 8.630 MHz ($L_1 = 100$ H).

The unit can be used to measure the Q of inductors. To that end, a potentiometer is connected in parallel with L1 and adjusted so that the current through the amplifier is doubled. The Q is then calculated from:

$$Q = \frac{R_p}{2\pi f\,L}$$

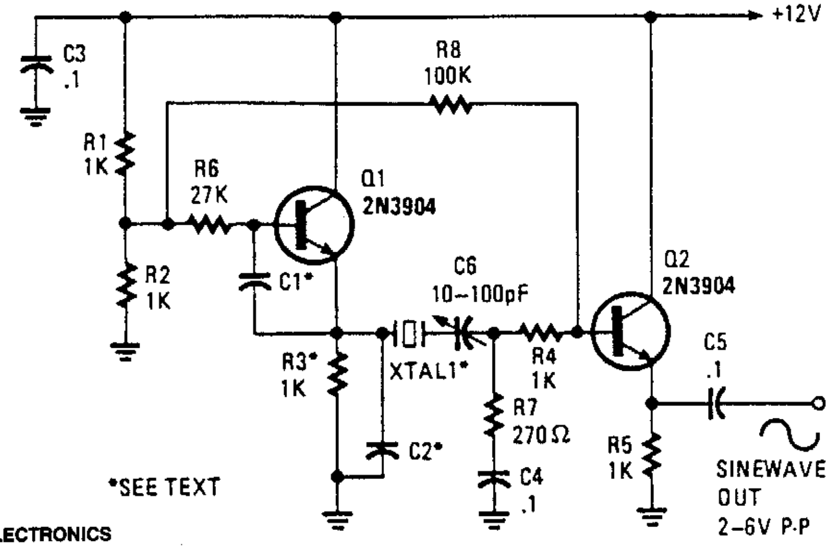
60-Hz SINE-WAVE GENERATOR



A chip by Micro Linear and two CMOS (40165 and 4060) chips generate sine waves at 60 Hz.

MICRO LINEAR *Fig. 79-4*

TWO-TRANSISTOR SINE-WAVE OSCILLATOR



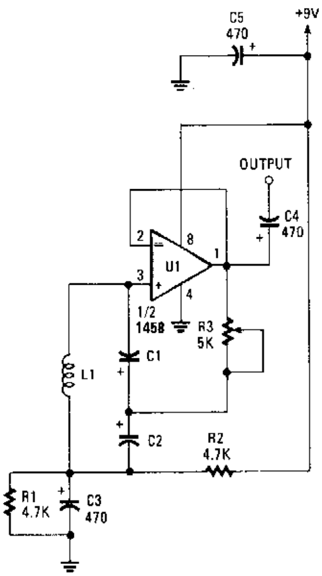
POPULAR ELECTRONICS

Fig. 79-5

This oscillator uses two transistors and operates the crystal in the fundamental mode. C_1 and C_2 should be about 2 700 pF for 1 MHz, 680 pF for 5 MHz, and 330 pF for 10 MHz. 150 pF can be used for up to 20 MHz. The output is a near perfect sine wave. Try varying C_1 and C_2 for best waveform. About 2 to 6 Vpp is available.

VLF AUDIO TONE GENERATOR

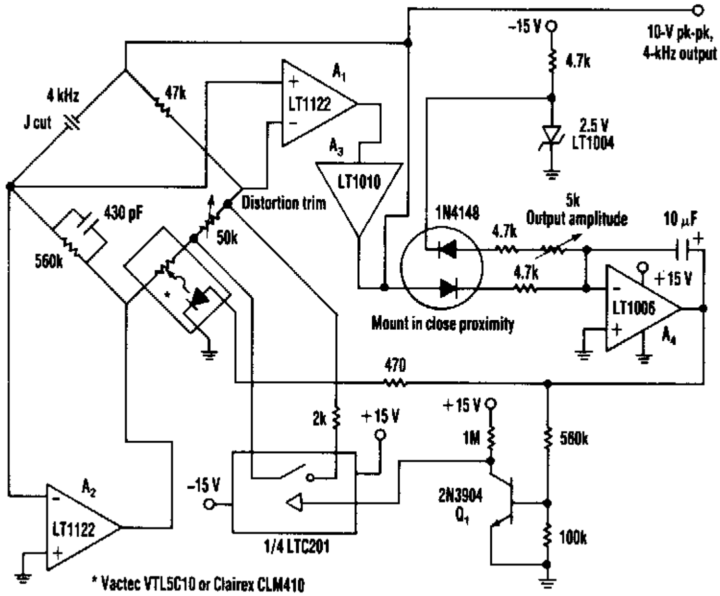
Using an LC-tuned circuit, this oscillator can produce frequencies of less than 10 Hz. C1 and C2 can be as large as 1 000 μF.



POPULAR ELECTRONICS

Fig. 79-6

VERY LOW DISTORTION OSCILLATOR



ELECTRONIC DESIGN

Fig. 79-7

This oscillator uses a bridge circuit with an optoisolator as a gain-control device. The resultant distortion can be held to 9 ppm (.000 9%) with proper adjustment.

LOW-FREQUENCY LC OSCILLATOR

POPULAR ELECTRONICS

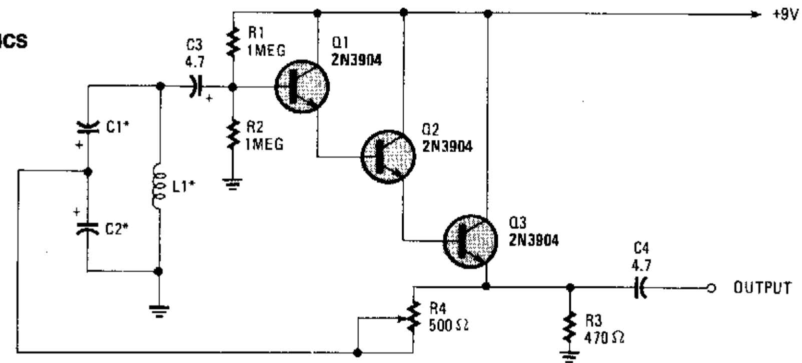


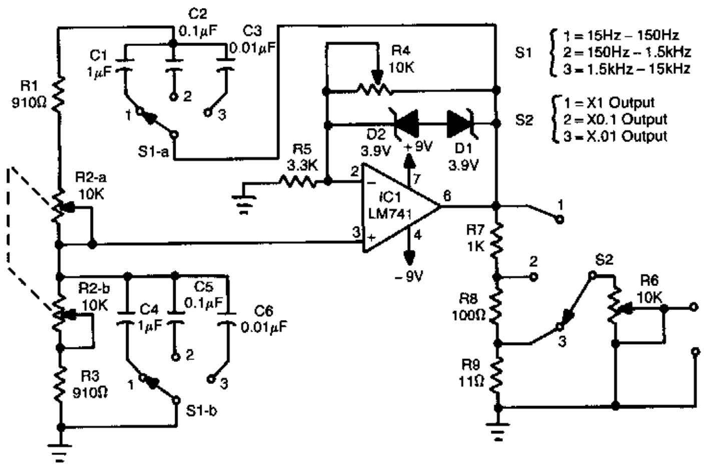
Fig. 79-8

Basically a Hartley oscillator using a triple-emitter follower, this oscillator can be used at audio and low radio frequencies. The frequency is given by:

$$f = \frac{1}{2\pi\sqrt{L_1 C_T}}, \text{ where } C_T = \frac{C_1 C_2}{C_1 + C_2}$$

At 1 kHz, typically C would be 4.7 μF tantalums, but this is only a guide as to convenient values to use.

THREE-DECADE 15-Hz TO 15-kHz WIEN-BRIDGE OSCILLATOR

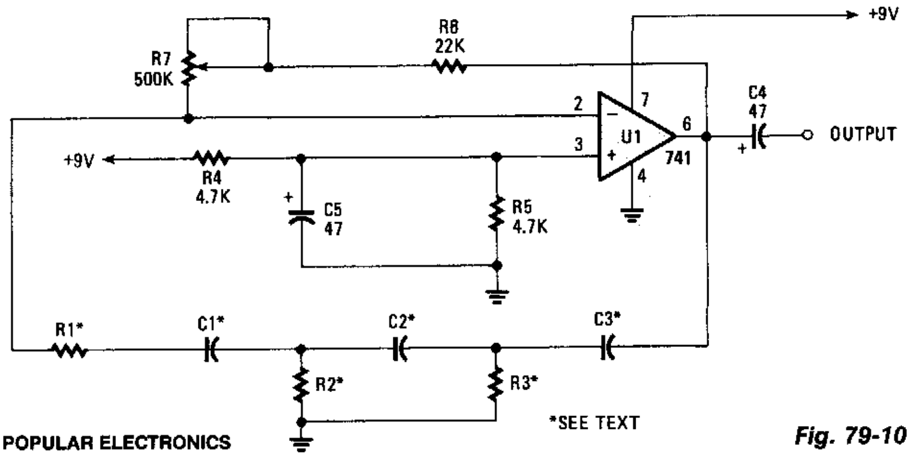


RADIO-ELECTRONICS

Fig. 79-9

In this circuit, an LM741 op amp drives a Wien-bridge network using two zener diodes as an amplitude limiter. Range selection is done by switch selecting the capacitors (C1 through C6) and tuning is done via a ganged pot. The output is about 8 Vpp max., depending on the setting of S2 and R6. R4 is set for maximum distortion consistent with stable output.

PHASE-SHIFT OSCILLATOR FOR AUDIO RANGE



POPULAR ELECTRONICS

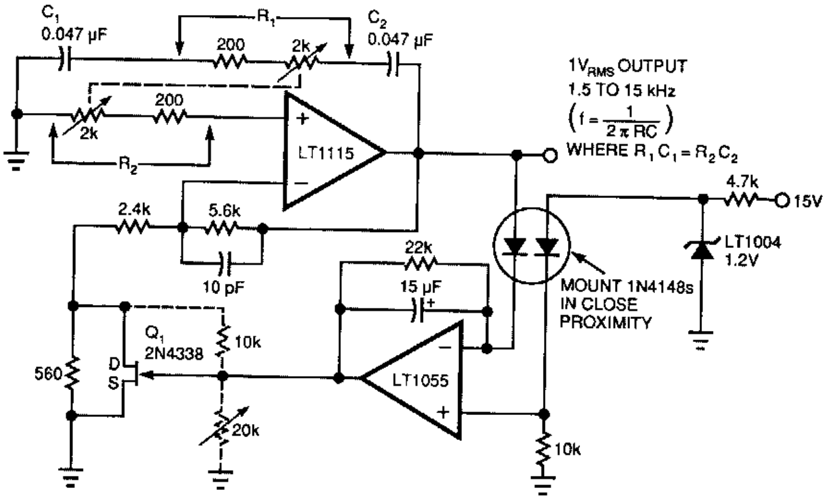
Fig. 79-10

This phase-shift oscillator is useful for audio oscillator applications. Adjust R7 for a good sine wave. An amplifier gain of 29 is required for oscillation. If $C = C_1 = C_2 = C_3$ and $R = R_1 = R_2 = R_3$:

$$f = \frac{1}{2\pi\sqrt{6}RC}$$

Typically, R will be 1 to 100 kΩ and C will be 0.1 μF down to 100 pF, in most practical circuits. As a start, choose $R = 10$ kΩ and $C = 0.006\,8$ μF (for ≈ 1-kHz range).

WIEN-BRIDGE OSCILLATOR

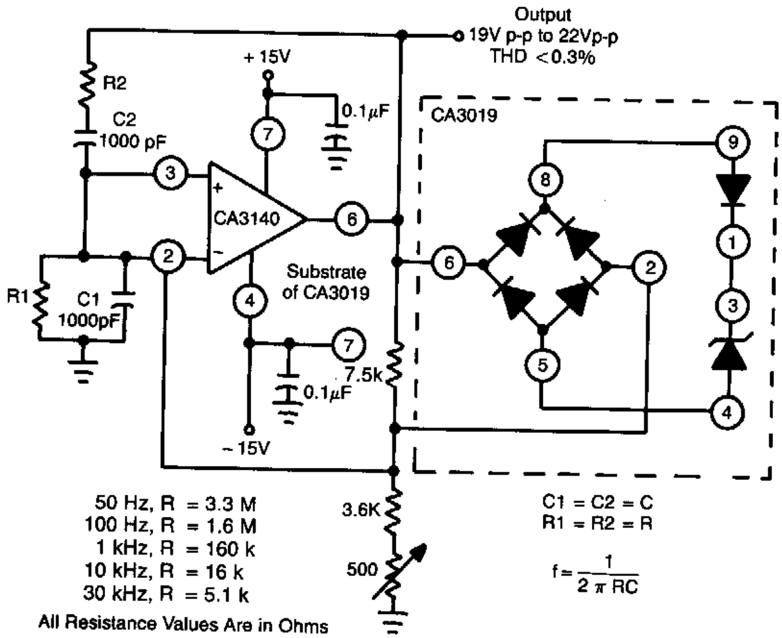


EDN

Fig. 79-11

This complex oscillator circuit uses a photocell and common-mode-suppression circuitry to achieve distortion of 0.0003%.

WIEN-BRIDGE OSCILLATOR

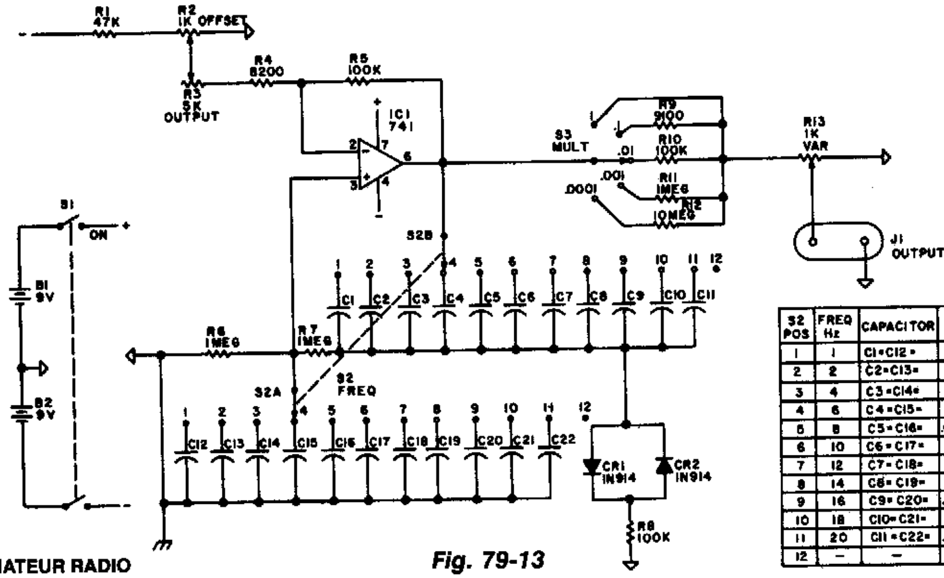


GE/RCA

Fig. 79-12

This circuit makes excellent use of high input impedance, high slew rate, and high-voltage qualities of CA3140 BiMOS op amp, in combination with CA3019 diode array.

LOW-FREQUENCY SINE-WAVE GENERATOR



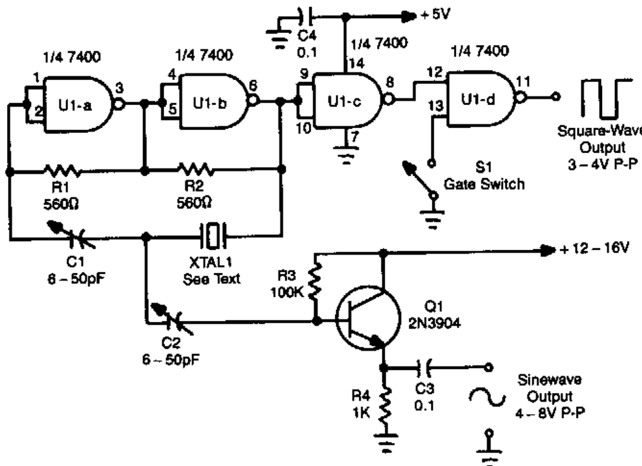
S2 POS	FREQ Hz	CAPACITOR	VALUE, µF
1	1	C1=C12=	.5+.1
2	2	C2=C13=	.15+.15
3	4	C3=C14=	.15
4	6	C4=C15=	.1
5	8	C5=C16=	.068+.0068
6	10	C6=C17=	.05+.01
7	12	C7=C18=	.05
8	14	C8=C19=	.033+.01
9	16	C9=C20=	.033+.0047
10	18	C10=C21=	.033
11	20	C11=C22=	.015+.015
12	—	—	—

73 AMATEUR RADIO

Fig. 79-13

Using a Wien-bridge oscillator, this circuit generates switch-selected frequencies from 1 to 20 Hz. A 741 op amp is used as the active element.

SINE- AND SQUARE-WAVE TTL OSCILLATOR

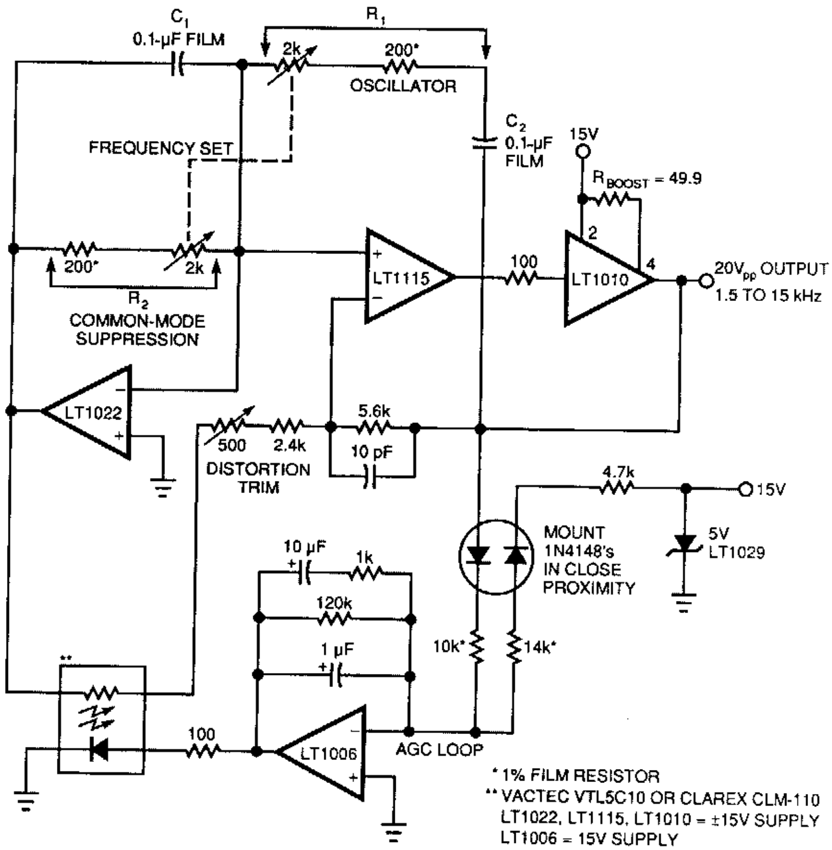


POPULAR ELECTRONICS

Fig. 79-14

Using a Quad NAND Gate, the TTL oscillator can use fundamental crystals between 1 and 10 MHz. The sine-wave output is taken directly from the crystal, which acts as its own filter, and yields a fairly clean sine wave. Adjust C1 and C2 for the best sine wave and also to set the crystal frequency. TTL square wave can be taken from U1D. The gate switch can be replaced with a logic gate to electronically control the output.

WIEN-BRIDGE-BASED OSCILLATOR WITH VERY LOW DISTORTION



EDN

Fig. 79-15

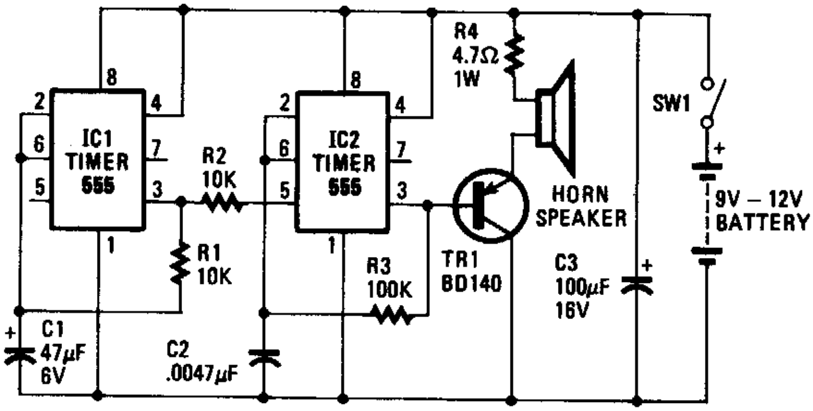
This complex oscillator circuit uses a photocell and common-mode-suppression circuitry to achieve distortion of 0.0003%. This oscillator circuit replaces the lamp in the traditional Wien bridge with an electronic equivalent.

80

Sirens, Warblers, Wailers, and Whoopers

The sources of the following circuits are contained in the Sources section, which begins on page 677. The figure number in the box of each circuit correlates to the entry in the Sources section.

"HEE-HAW" ELECTRONIC SIREN



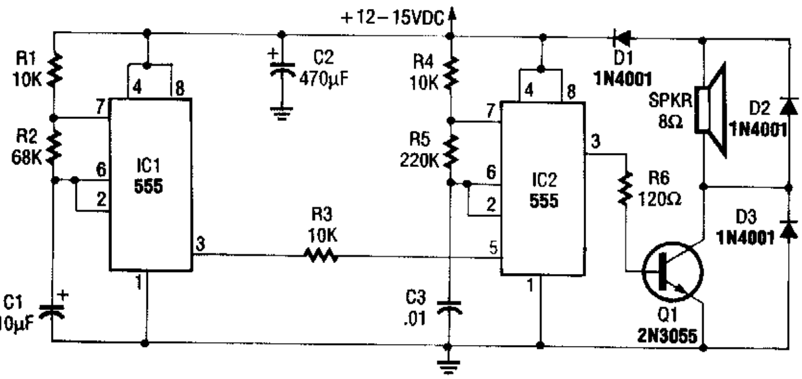
HANDS-ON ELECTRONICS

Fig. 80-1

The oscillator based on IC2 is responsible for producing the sound. Its output is connected to the base of TR1, which amplifies it to drive the speaker. Resistor R4 is included in the circuit to limit the current through TR1 to a safe and reasonable level. The oscillation frequency of IC2 is partially dependent on the values of R_3 and C_2. Another factor that governs the frequency of oscillation is the magnitude of voltage fed to pin 5 of IC2. If a voltage of varying magnitude is fed to pin 5, the internal circuitry of the IC is forced to reset at a different rate, which changes the frequency.

IC1 is also connected as an oscillator,but it runs much slower than IC2: around 1 Hz. Each time the IC triggers, the voltage at pin 3 goes high. As pin 3 is connected to pin 5 of IC2, this forces IC2 to change its note. That produces the "hee-haw" sound of the siren.

ALTERNATE WARBLE-TONE SIREN

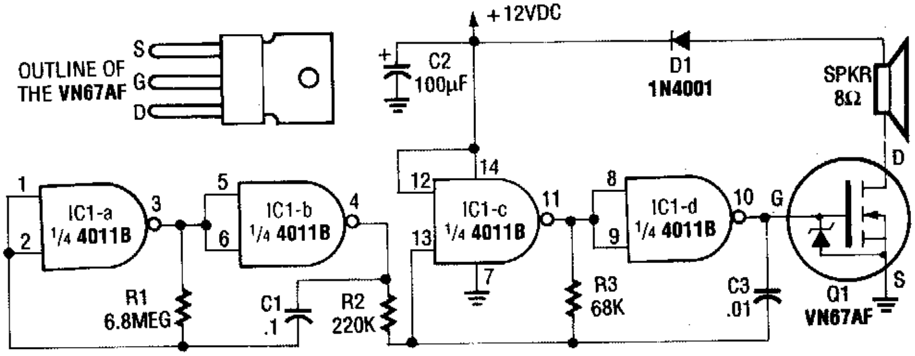


RADIO-ELECTRONICS

Fig. 80-2

This circuit uses two NE555 timers to generate a warble tone. IC1 frequency shifts IC2 by feeding a square wave to pin 5, the modulation input of IC2. IC1 runs at about 1 Hz.

6-W WARBLE-TONE SIREN

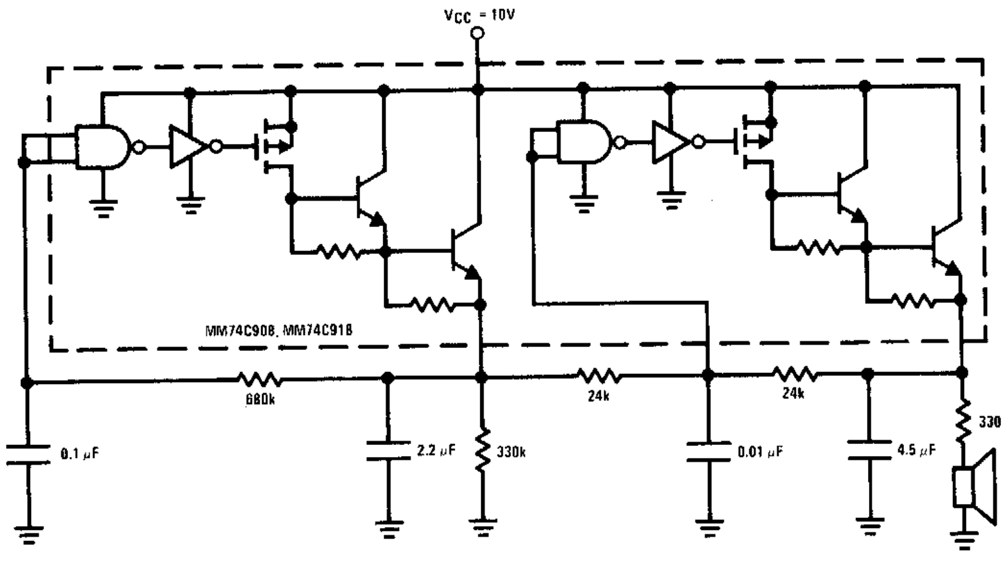


RADIO-ELECTRONICS

Fig. 80-3

This circuit uses a CMOS chip and a VMOS FET amplifier for 6 W of audio output. 18 W of audio can be generated using a +24-Vdc supply. IC1A and IC1B are used as a 1-Hz oscillator. IC1C and IC1D form a 1-kHz multivibrator that is gated by the 1-Hz signal from IC1A and IC1B.

LOW-COST SIREN

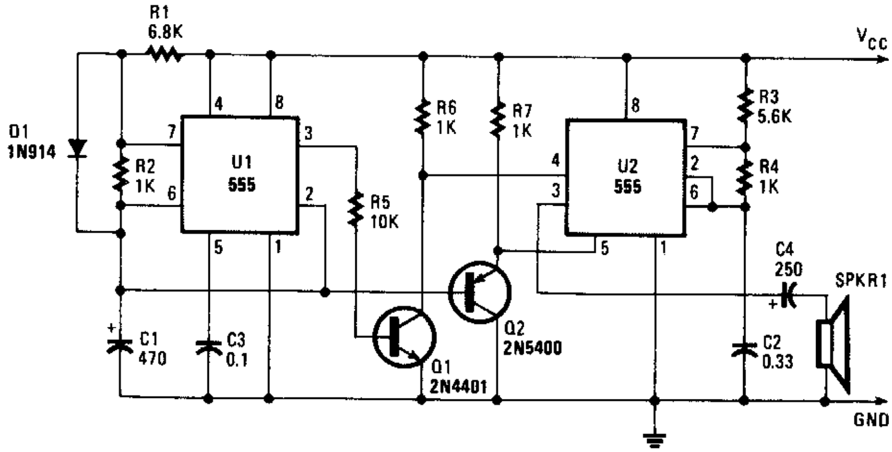


NATIONAL SEMICONDUCTOR

Fig. 80-4

This low-cost 1-package siren has one VCO, and the other oscillator generates the voltage ramp to vary the frequency at the VCO output. All components within the dotted line are part of the IC.

WHOOPER



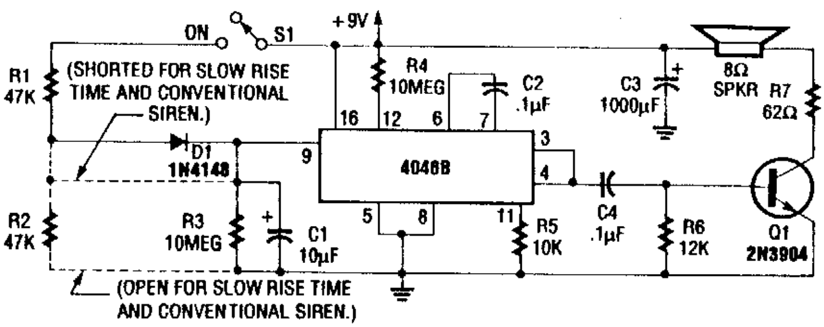
POPULAR ELECTRONICS

Fig. 80-5

Integrated circuit U1 is connected as a low-frequency asymmetrical oscillator. Its output is inverted by Q1 and fed to the reset terminal of U2 at pin 5. Integrated circuit U2 is configured as an audio oscillator and is enabled when the output of U1 is low. With the voltage at pin 5 of U2 constant, the circuit just ''bleeps.''

The voltage across capacitor C1 is fed to the base of Q2, which turns it on and grounds pin 5 of U2. When the frequency of the reset signal on pin 4 falls, the output frequency of U2 rises. The output then becomes a whoop, starting low in frequency and ending high. Resistor R1 sets the repetition rate and R2 determines the time duration of the whoop. Resistors R3 and R4 set the center-operating frequency.

ELECTRONIC SIREN



RADIO-ELECTRONICS

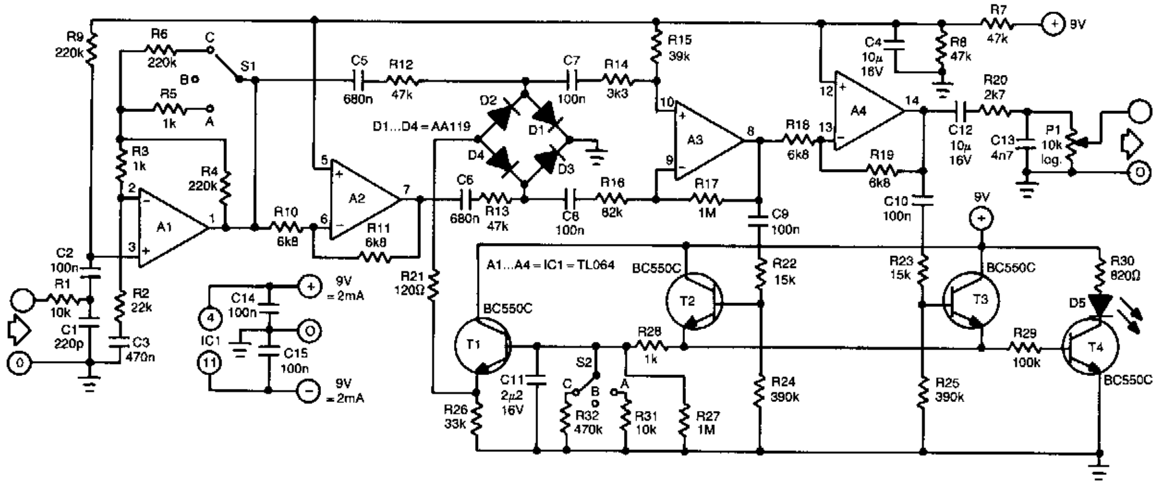
Fig. 80-6

For normal wailing tone, short D1 and open R2. For fast rise and slow fall in frequency, include D1 and R2. Use of a CD4046B with a diode-RC network as shown produces a siren tone, using a VCO.

81

Sound Effects

The sources of the following circuits are contained in the Sources section, which begins on page 677. The figure number in the box of each circuit correlates to the entry in the Sources section.



ELEKTOR ELECTRONICS *Fig. 81-1*

The control of this compressor is based on the dependence of the dynamic resistance of a diode on the current flowing through it. The heart of the present circuit is the diode bridge (D1 through D4), which behaves as a variable resistance controlled by the current flowing in T1.

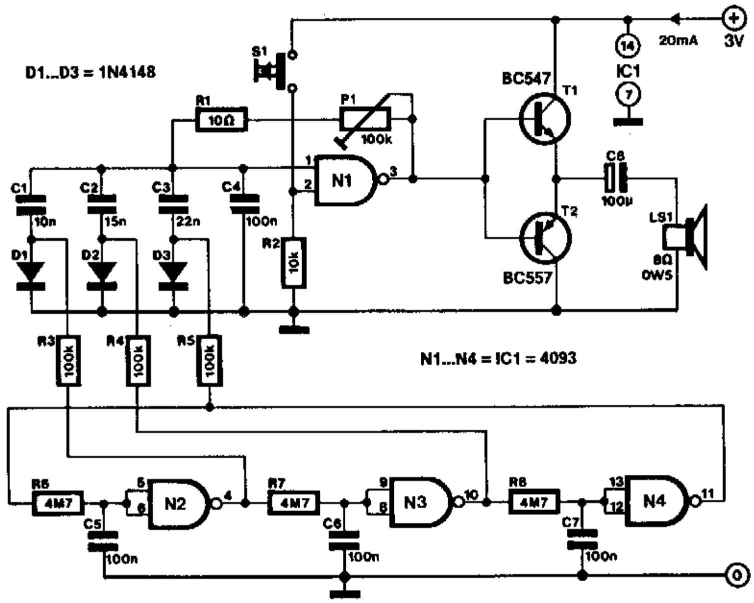
The input signal is applied to preamplifier state A1 via low-pass filter R1/C1 that removes any HF noise from the input.

Switch S1 in the feedback loop of A1 sets the amplification to 1 (position A), 6 (C) or 11 (B). The amplified signal is applied directly to the diode bridge via R12 and C5, and inverted via inverter A2, capacitor C6 and resistor R13.

The two signals are summed by the bridge, amplified (in A3), then split again into two, one of which is inverted by A4. The positive half cycles of the two signals are used to switch on T2 and T3, respectively. Capacitor C11 is then charged via R12. When the potential across this capacitor reaches a certain level, T1 is also switched on, after which a control current flows through the bridge via R21. This current lowers the resistance of the bridge so that the signal is attenuated (compressed). At the same time, the LED lights to indicate that the signal is being compressed. Capacitor C12 prevents any dc voltage from reaching the output.

The output signal is taken from the wiper of P1. Low-pass section R20/C13 limits its bandwidth to 12 kHz. Switch S2 enables the selection of various decay times from C11. The values shown in the diagram are the most useful. Nevertheless, these values are subjective and can be altered to personal taste and requirements.

SINGLE-CHIP MELODY GENERATOR

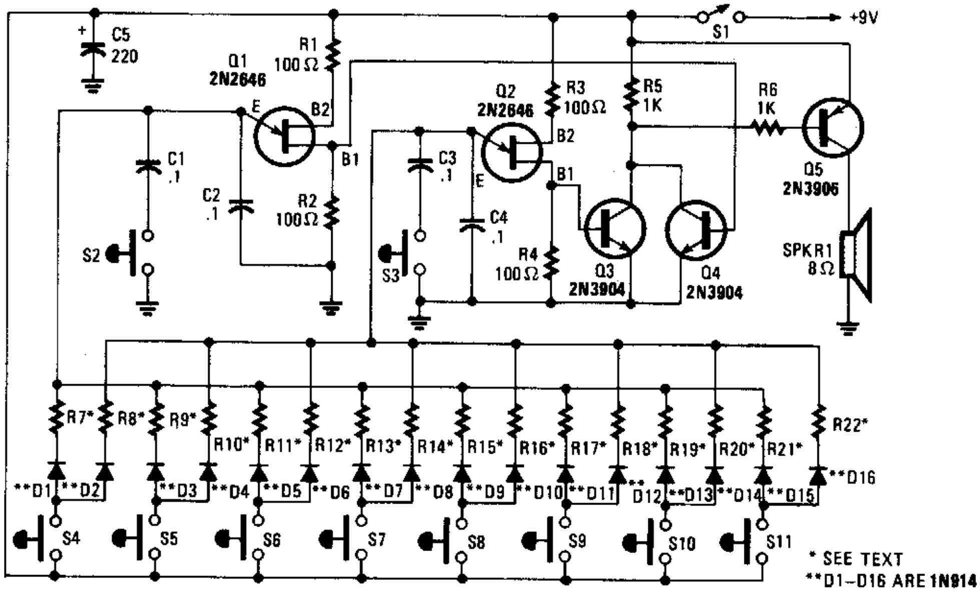


ELEKTOR ELECTRONICS *Fig. 81-2*

This melody generator, based on a 4093 CMOS Schmitt trigger, can be used in alarms, doorbells, and cars (audible reverse gear or lights-on indicator).

Three of the four NAND gates in the 4093 are connected in series by RC networks. Oscillation is affected by feedback of the output signal of N4 to the input of N2. The logic-high levels produced by the cascaded gates in the oscillator circuit are used to bias one of associated diodes D1, D2, or D3). The relevant diode connects one of the frequency-determining capacitors (C1 through C3) to tone oscillator N1. The audio signal available when S1 is pressed is applied to complementary transistor pair (T1/T2) that drives the loudspeaker. The frequency of the emitted tone can be adjusted to individual taste with preset P1.

ELECTRONIC BAGPIPE

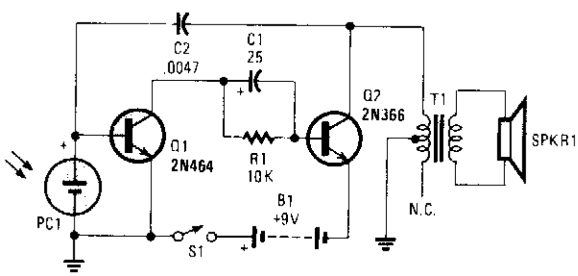


POPULAR ELECTRONICS

Fig. 81-3

The electronic bagpipe mimics the sound of real instruments. This circuit uses two UJT oscillators and an amplifier (Q3, Q4, and Q5). R7 through R22 are selected for tonal range desired (typically 3 300 Ω). Each key selects resistors for the two oscillator circuits Q1 and Q2. S2 and S3 vary the tonal range of S4 through S11.

ELECTRONIC MUSIC MAKER

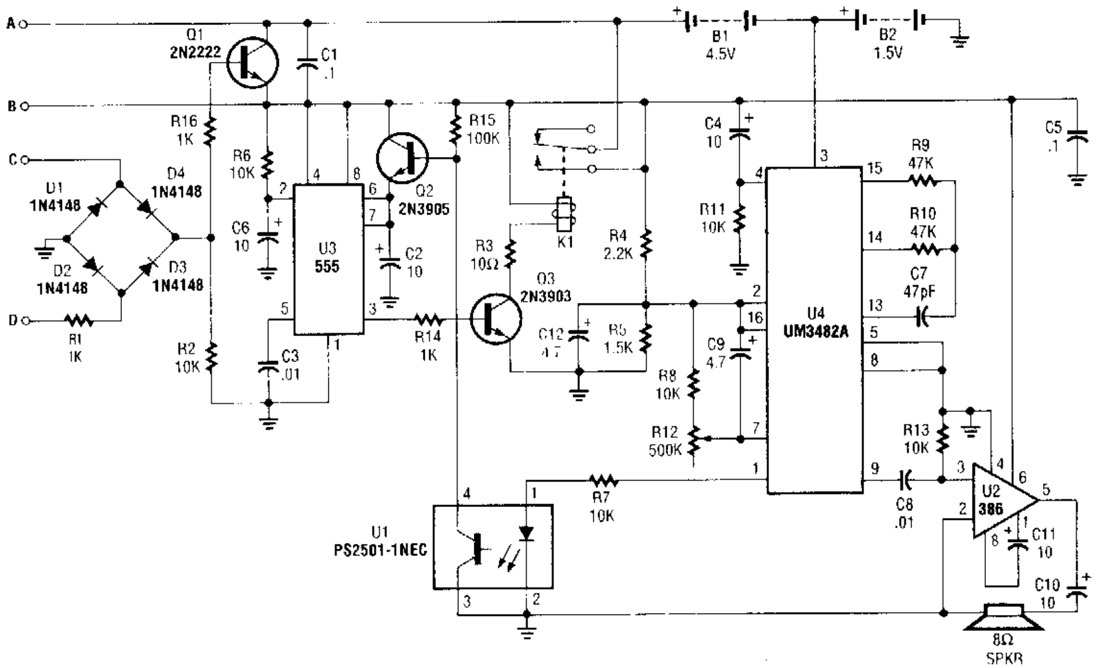


POPULAR ELECTRONICS

Fig. 81-4

This electronic music maker uses an astable oscillator circuit that is controlled by a photocell. The light falling on the photo cell controls the tone. By mounting the circuit in a box, you can control light-reading PC1 with your hand.

MUSICAL DOORBELL

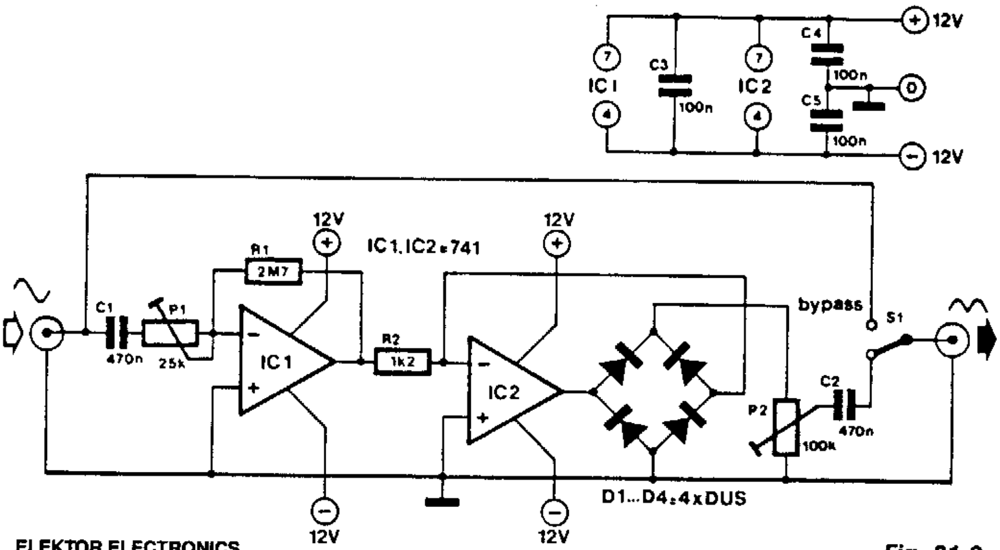


POPULAR ELECTRONICS

Fig. 81-5

8 to 15 Vac is applied to terminals C and D, which produces a dc voltage across R2, and turns on Q1. This connects the batteries B1 and B2 to the rest of the circuit, which activates it. Latch U3 is triggered, it remains on until Q2 turns on, charges C2, and turns off U2. When U3 is turned on, Q3 is forward-biased, which energizes K1, powers up U4. At the time the K1 contacts close, C4 couples a positive spike to pin 4 of U4, a melody synthesizer chip. U4 generates a pre-programmed tune, at the end of which pin 1 of U4 goes positive. This activates optocoupler U1 and turns off Q2, which drops out the relay. U2 acts as an audio amplifier, which drives an 8-Ω speaker.

OCTAVE SHIFTER FOR MUSICAL EFFECTS

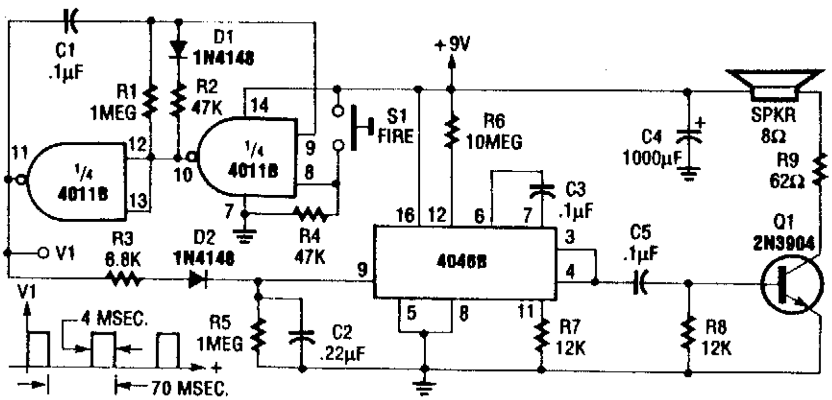


ELEKTOR ELECTRONICS

Fig. 81-6

This musical special-effect device is basically a frequency doubler. The input audio is amplified and doubled by using a full-wave rectifier, which has a dc output plus a twice the ac frequency component. The 2× frequency component is fed to the output.

PHASOR SOUND GENERATOR

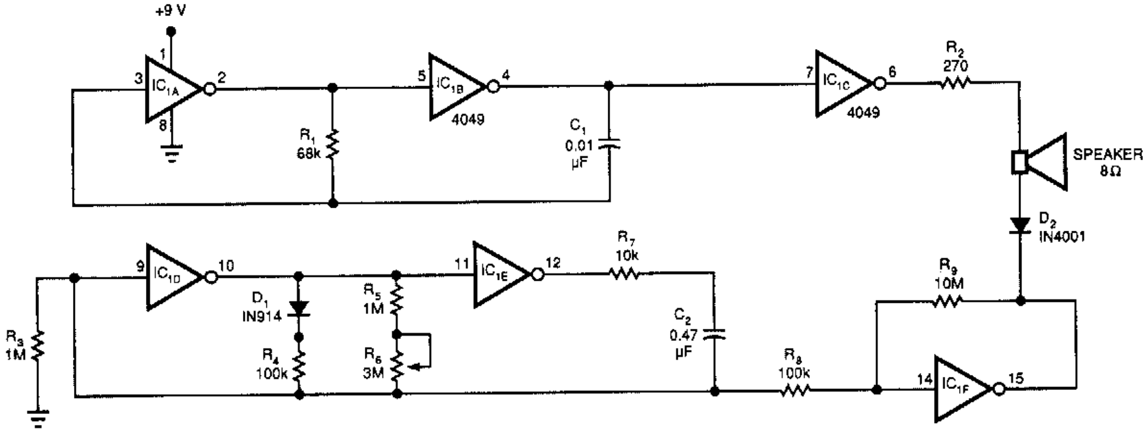


RADIO-ELECTRONICS

Fig. 81-7

The 4011B astable is gated by S1 to produce 4-ms pulses at 70-ms intervals. Each pulse charges C2 via D2, producing a high tone that decays slowly as C2 discharges through R5. The process repeats for each pulse.

SINGLE-CHIP CHIME



NOTES: 1. IC1 = 4049.
2. ALL RESISTORS ARE 1/4 W, 10%.
3. ALL CAPACITORS ARE 25 VDC, 20%.

EDN

Fig. 81-8

This circuit uses only one IC, produces a pleasant tone, and sports a single control for adjusting the tone's chiming rate. IC1A and IC1B form an astable multivibrator, which produces the circuit's basic tone. The multivibrator's frequency is:

$$f = \frac{1}{2.2 \times R_1 \times C_1}$$

The component values produce a 668-Hz tone. IC1C buffers the multivibrator's output to the 8-Ω speaker. Current-limiting resistor R2, determines the speaker's volume. R_2 minimum value is 220 Ω. IC1D and IC1E form an asymmetric, astable multivibrator, which adds a chime effect to the circuit's basic tone. The chime effect's frequency is:

$$t_{LO} = 1.1C_2(R_4 \parallel (R_5 + R_6))$$

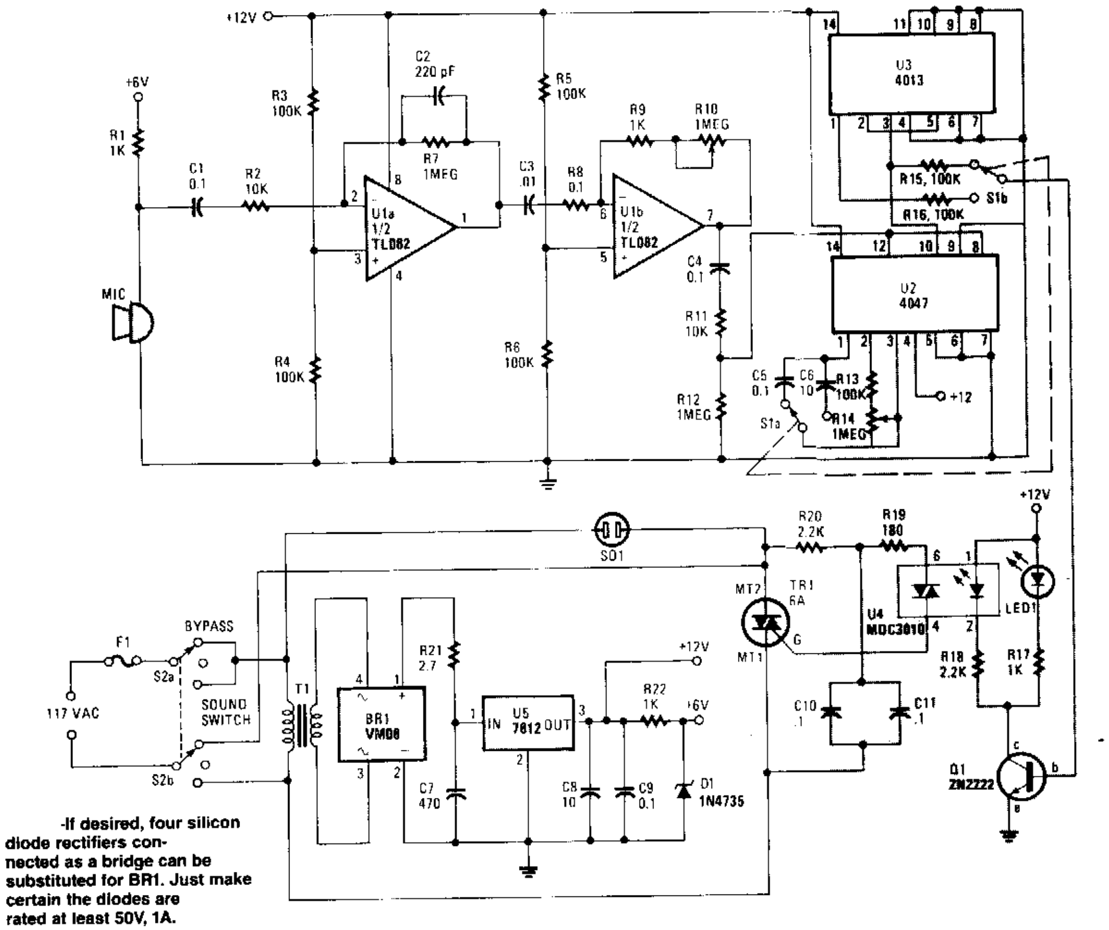
$$t_{HI} = 1.1C_2(R_5 + R_6)$$

R_7 gives this rate multivibrator a slowly varying output signal to produce a pleasant decay for the chime effect. IC1F is an inverting amplifier for the chime multivibrator.

82

Sound Operated Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 678. The figure number in the box of each circuit correlates to the entry in the Sources section.

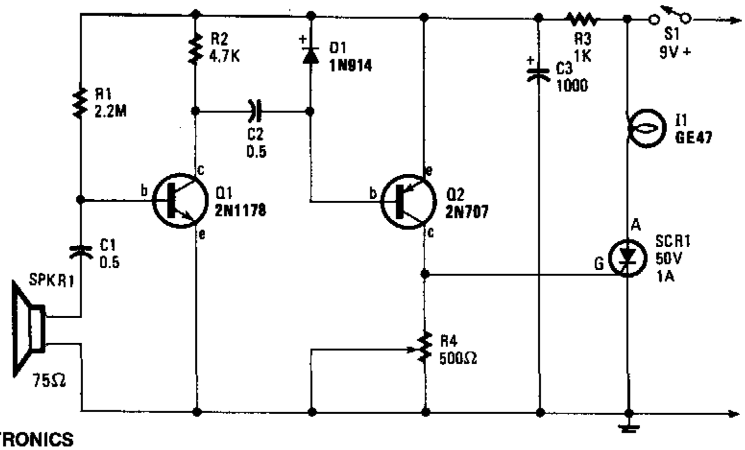


RADIO-ELECTRONICS

Fig. 82-1

This circuit provides either latched switching or timed switching. U1A and U1B provide audio amplification from the microphone. U2 is a retriggerable monostable multivibrator. S1A and S1B select either U3, a flip-flop, or U2. R13 and R14 allow a 6- to 60-second timer delay after the sound ceases, in the timed mode. BR1, U5, and associated components form a power supply. Q1 drives optocoupler U4 and triggers triac TR1.

SOUND-OPERATED SWITCH

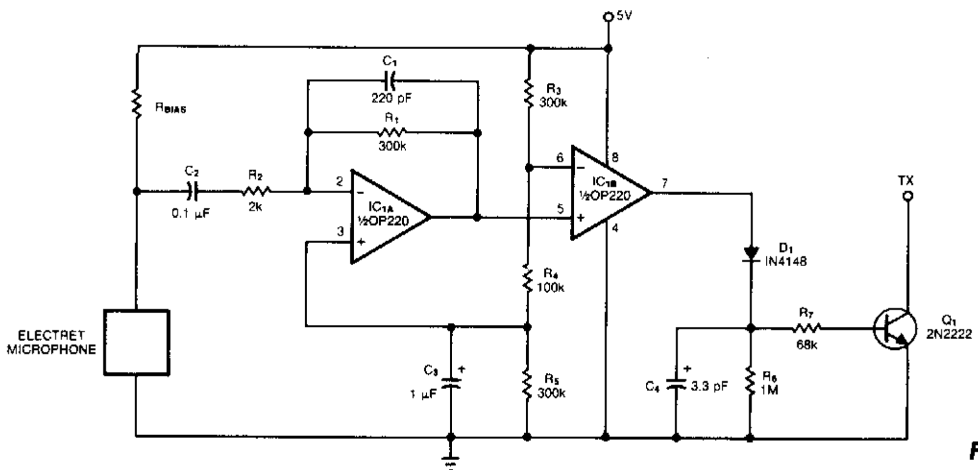


HANDS-ON ELECTRONICS *Fig. 82-2*

This sound-operated switch will sense the ring of the phone and translate this to a lamp that will go on and off. The amplified signal across R2 reaches D1 through capacitor C2. The rectified audio signals provide a negative bias for Q2, a pnp transistor. This causes Q2 to conduct so the current that triggers SCR1 is provided at the gate. Potentiometer R4 sets the sensitivity. R3 and C3 delay the operating voltage for Q1 so that the circuit will not be triggered on by the sound of the on/off switch, S1 or by the current surge.

Set the lamp atop a TV receiver, turn it on and set the potentiometer so that a finger snap at two feet will trigger the lamp on. Place the speaker close to the telephone and give it a try.

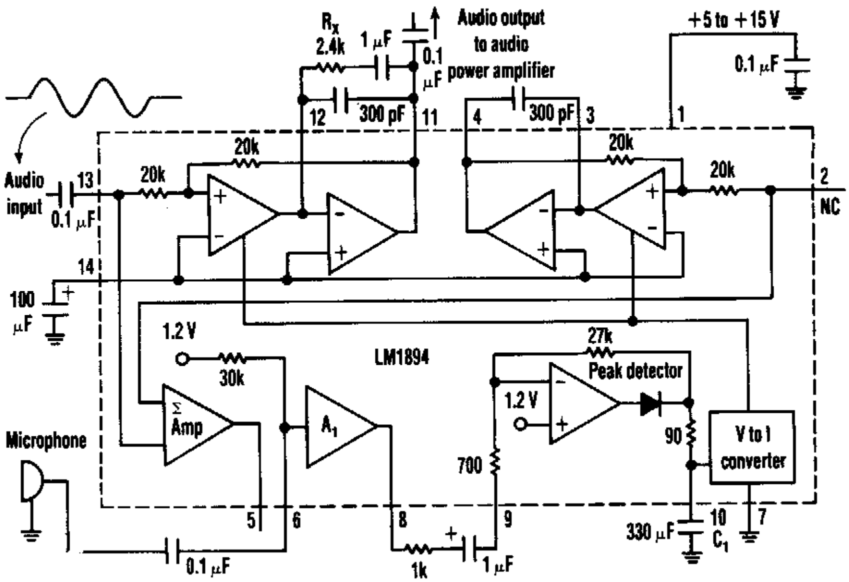
MICROPHONE-CONTROLLED VOICE-ACTIVATED SWITCH



EDN *Fig. 82-3*

An electret microphone feeds a bandpass filter circuit (IC1A), then feeds a comparator, which in turn drives Q1. Q1 is a switch that conducts when audio from IC1B causes D1, C4, R6, and R7 to bias it ON.

GAIN-CONTROLLED AMPLIFIER



ELECTRONIC DESIGN *Fig. 82-4*

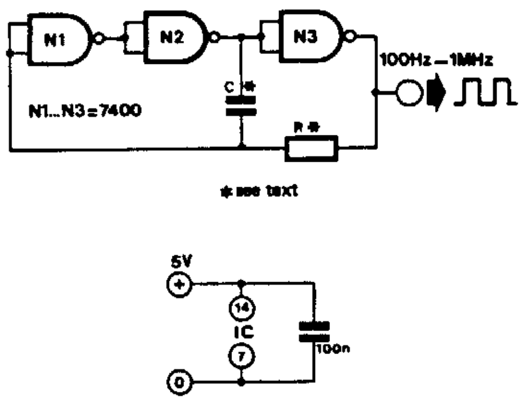
This single-chip circuit adjusts its audio gain according to the ambient noise picked up by the microphone. When operating in a quiet environment, the audio output is quiet, while a noisy environment results in a louder audio output. Audio to pin 13 is amplified by the variable-gain amplifier within the LM1894 IC. Audio from the microphone connected through 0.1-μF capacitor to pin 6 controls the audio gain of the variable-gain amplifier. The output appears on pin 11 and is taken off through an 0.1-μF capacitor.

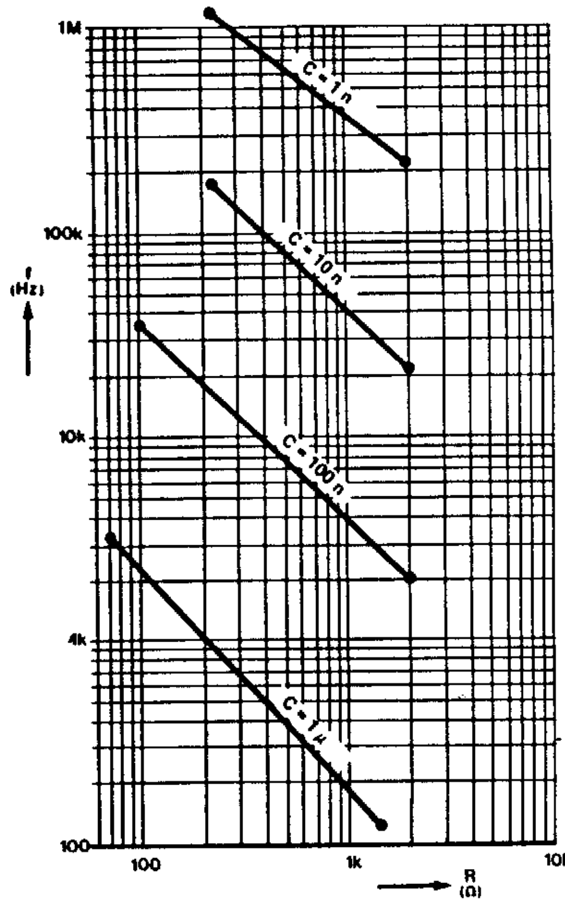
83

Square-Wave Generators

The sources of the following circuits are contained in the Sources section, which begins on page 678. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE TTL, LSTTL, CMOS, SQUARE-WAVE GENERATORS

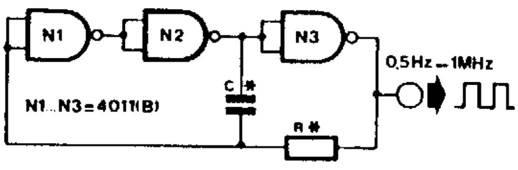


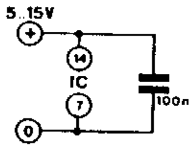


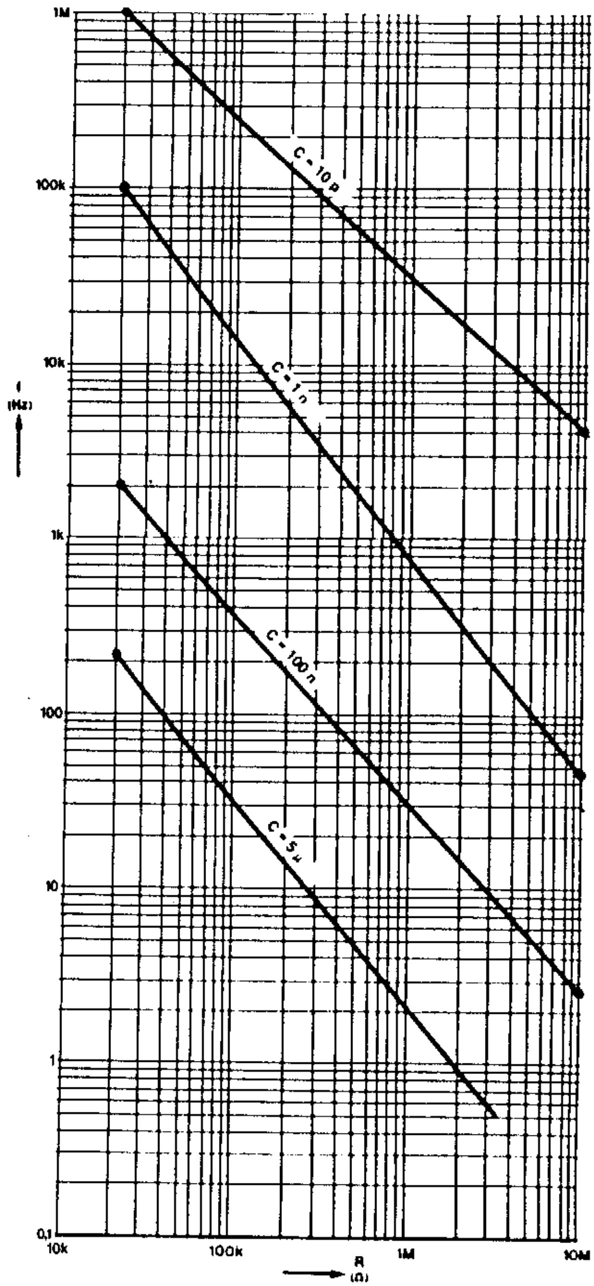
ELEKTOR ELECTRONICS

Fig. 83-1(a)

These three circuits for TTL, LSTTL, and CMOS logic use three gates each, and one or two resistors and capacitor as a square-wave oscillator. The circuits are useful for clock oscillators, etc. *R* and *C* are determined from the nomographs.

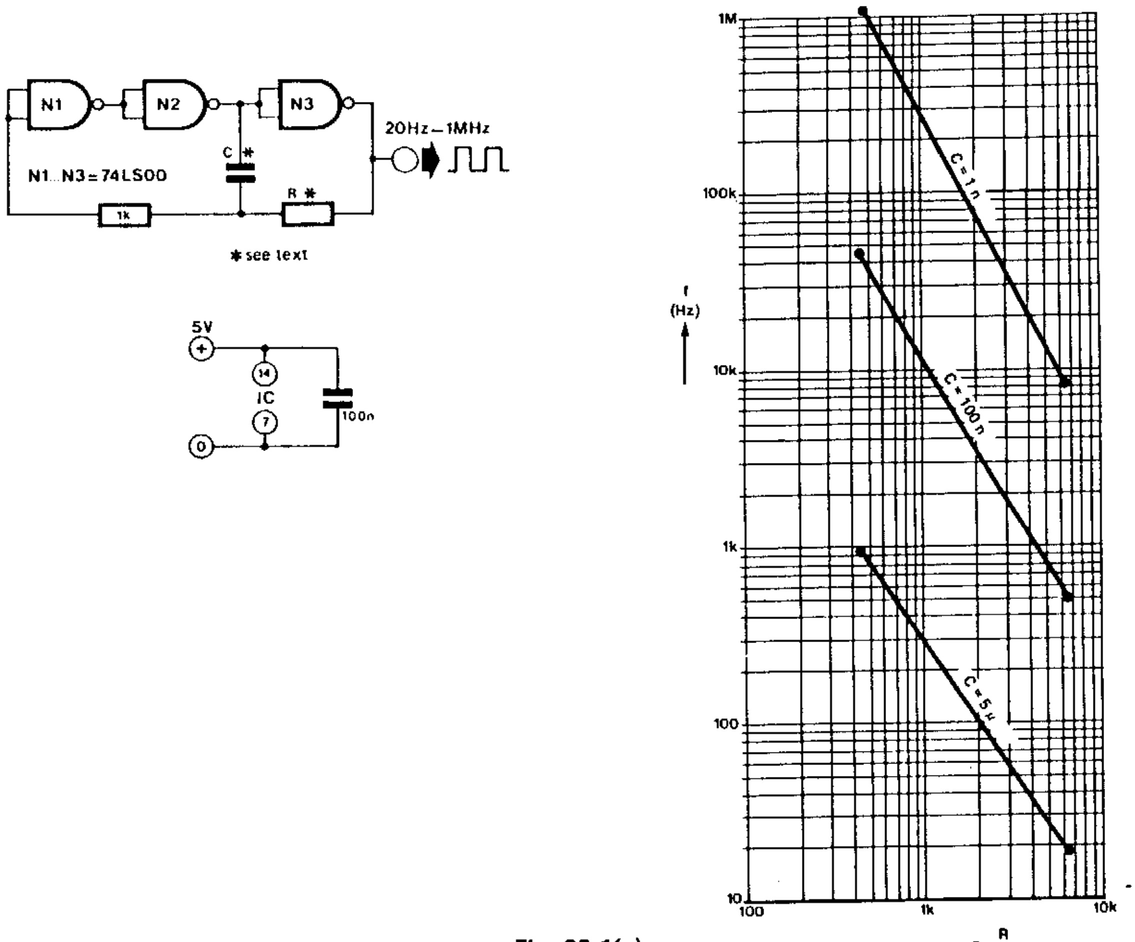






ELEKTOR ELECTRONICS

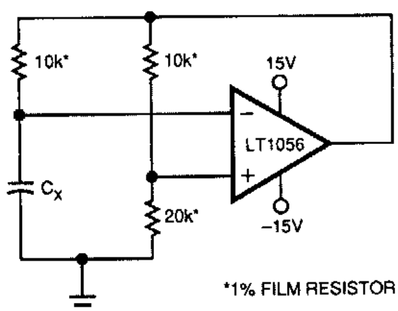
Fig. 83-1(b)



ELEKTOR ELECTRONICS

Fig. 83-1(c)

SIMPLE SQUARE-WAVE OSCILLATOR

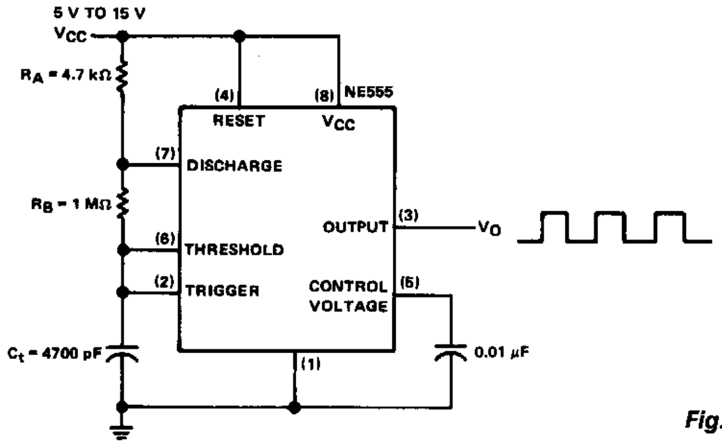


EDN

Fig. 83-2

Using only four components, this circuit generates a square wave. Oscillation frequency is ≈ $1/RC_X$ Hz, $R = \text{M}\Omega$, $C_X = \mu\text{F}$ (in this case, $R = 10\ \text{k}\Omega$).

SQUARE-WAVE OSCILLATOR



TEXAS INSTRUMENTS

Fig. 83-3

The NE555 is connected in the astable mode and uses only three timing components (RA, RB, and Ct. A 0.01-μF bypass capacitor is used on pin 5 for noise immunity. The operating restrictions of the a-stable mode are few. The upper frequency limit is about 100 kHz for reliable operation, as a result of internal storage times. Theoretically, it has no lower frequency limit, only that which is imposed by R_t and C_t limitations. The frequency for the circuit can be calculated as:

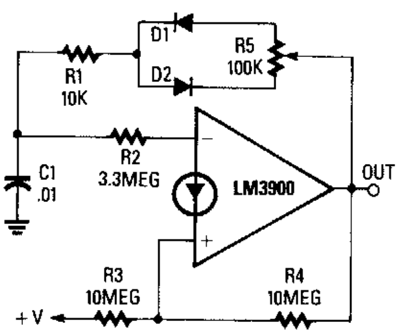
$$f = \frac{1.44}{(R_A + 2R_B)C_t}$$

$$= \frac{1.44}{(4.7\ \text{k}\Omega + 2\ \text{M}\Omega)\ (0.0047\ \mu\text{F})}$$

$$= \frac{1.44}{9.42209 \times 10^{-3}}$$

$$f = 152.8\ \text{Hz}$$

VARIABLE DUTY-CYCLE SQUARE-WAVE GENERATOR

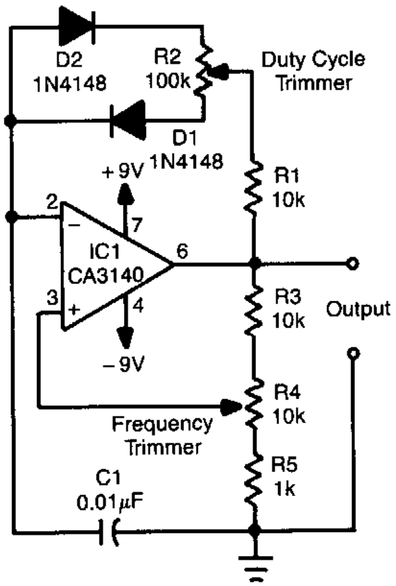


RADIO-ELECTRONICS

Fig. 83-4

C1 alternately charges via R1/D1 and the upper half of R5, and discharges via R1/D2 and the lower half of R5. The duty cycle can be varied over the range from 1:10 to 10:1 via R5.

SQUARE-WAVE GENERATOR

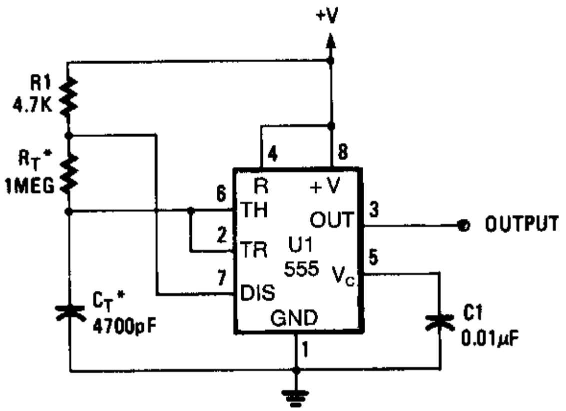


RADIO-ELECTRONICS

Fig. 83-5

This relaxation oscillator circuit uses diodes to produce charge and discharge paths for C1. The duty cycle is set via R2 and the frequency via R4. C1 can be varied to vary the frequency range, which, for this circuit is approximately 300 to 3 000 Hz.

SQUARE-WAVE ASTABLE CIRCUIT



*WITH $R1 < R_T$

$T = 1.368 R_T C_T$

$f_0 = \frac{0.722}{R_T C_T}$

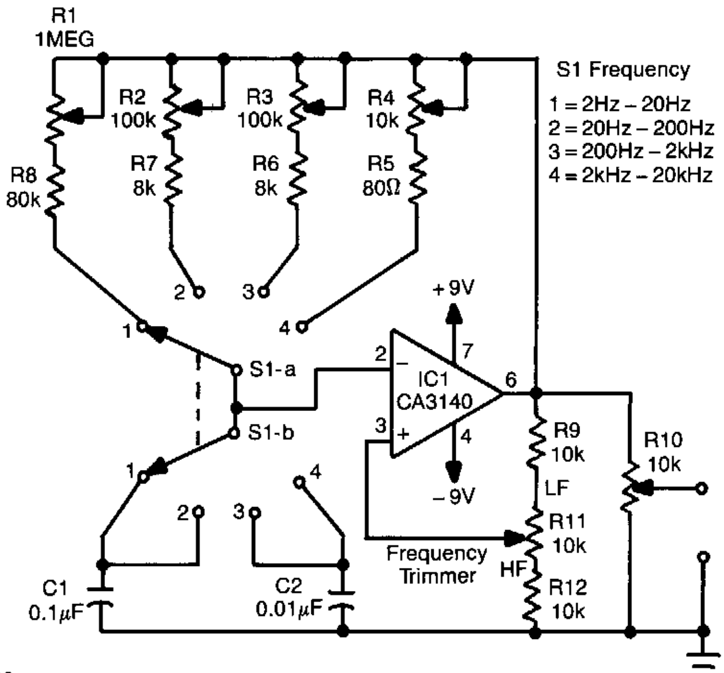
(AS SHOWN, $f_0 \cong 158$ Hz)

POPULAR ELECTRONICS

Fig. 83-6

This 555 circuit produces a square wave. The frequency depends on the values of R_T and C_T, as per the design equations.

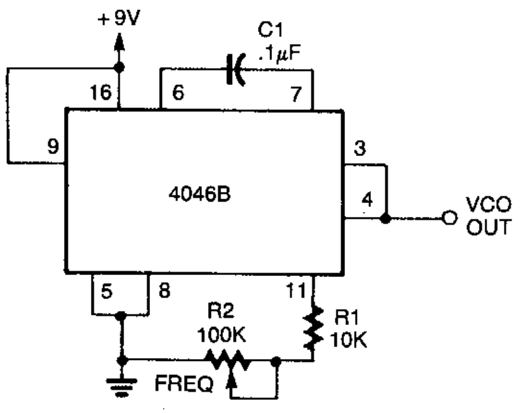
4-DECADE SQUARE-WAVE GENERATOR



RADIO-ELECTRONICS *Fig 83-7*

This circuit will generate a square wave of 2 Hz to 20 kHz. The circuit uses an op amp in a relaxation-oscillator configuration. The output is about 15 Vpp. R1 through R4 are calibration controls for each of the four frequency ranges, as selected by S1-a and S1-b. R10 adjusts the output level.

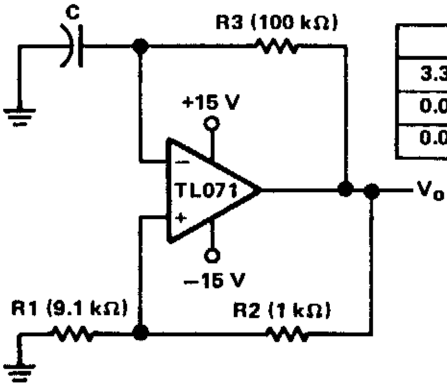
SIMPLE VARIABLE-FREQUENCY SQUARE-WAVE GENERATOR



A CD4046B PLL is used as a simple generator. The range is 200 Hz to 2 kHz, but it can be changed by changing C1.

RADIO-ELECTRONICS *Fig. 83-8*

BASIC MULTIVIBRATOR



C	f
3.3 μF	0.5 Hz
0.015 μF	108 Hz
0.0015 μF	1.0 kHz

TEXAS INSTRUMENTS

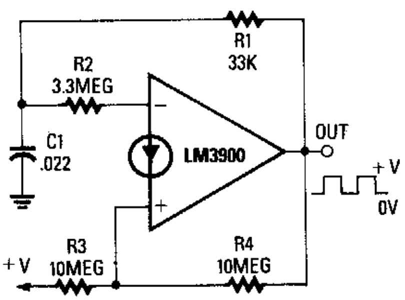
Fig. 83-9

When this circuit is turned on, the natural offset of the devices serves as an automatic starting voltage. Assume that output voltage V_O goes positive and the positive feedback through R2 and R1 forces the output to saturate. The high-voltage level at V_O, then charges C through R3, until the voltage at the inverting input exceeds that at the noninverting input.

As the inverting input exceeds the noninverting input level, the output switches to the negative saturation voltage. This action starts the capacitor discharging toward the new noninverting input level. When the capacitor reaches that level, the op amp switches back to the positive saturation voltage, and the process starts again. With the TL071, the positive and negative output levels are nearly equal, which results in a 50% duty cycle. The total time period of one cycle will be:

$$t_T = 2\ (R_3) C\ \mathrm{In}\left(\frac{1 + 2R_1}{R_2}\right)$$

1-kHz SQUARE-WAVE GENERATOR



RADIO-ELECTRONICS

Fig. 83-10

When the output is high, R3 and R4 are in parallel, and C1 charges via R1 until the current in R2 equals that at the noninverting terminal. This action occurs when C1's voltage rises to 2/3 of the supply voltage. At that point, the circuit switches regeneratively. The output switches low and C1 starts to discharge via R1.

Now, R4 is effectively disabled and the current to the noninverting terminal is determined solely by R3, so C1 discharges until the current through R2 falls slightly below that of R3.

This happens when the voltage across C1 falls to about 1/3 of the supply voltage. At that point, the circuit again switches regeneratively, and the output again goes high.

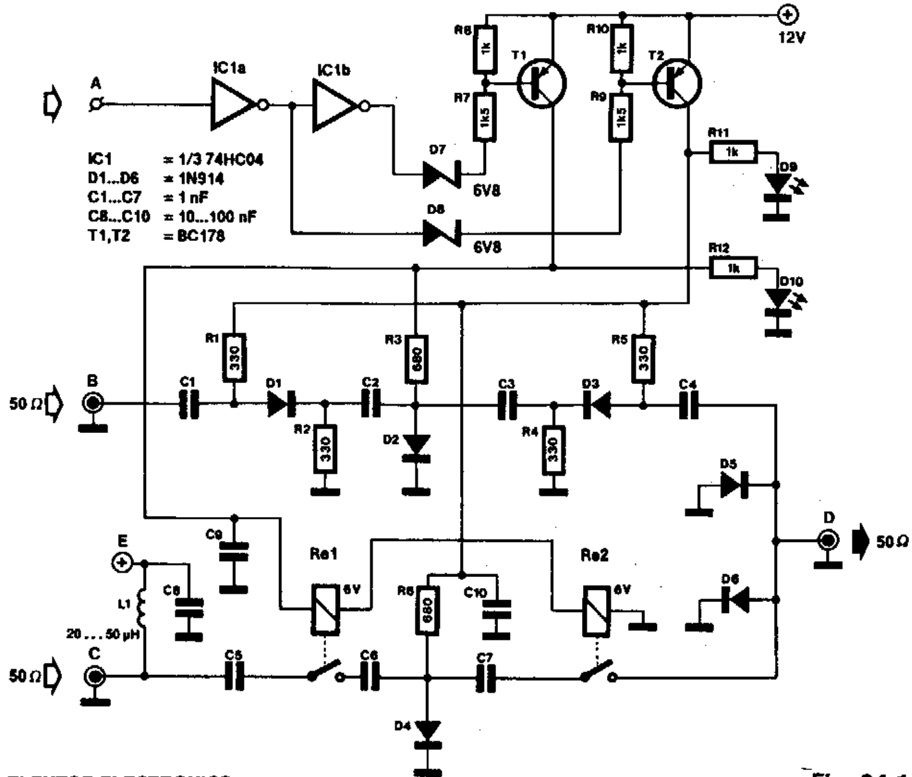
This circuit is useful for generating symmetrical square waves with maximum frequencies of only a few kHz. Because of the poor slew-rate characteristics of the LM3900 (0.5 V/μs), the output waveforms have rather slow rise and fall times.

84

Switching Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 678. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC ANTENNA SELECTOR



ELEKTOR ELECTRONICS

Fig. 84-1

The electronic antenna selector is intended to switch between two FM antennas via a logic signal. Gates IC1A and IC1B ensure a clean switching action and at the same time form the interface between the 5-V logic level (probably available from the receiver) and the 12-V supply voltage for the selector. Depending on the type of gate used, a digital TTL or CMOS control signal is available in direct and inverted form at the outputs of IC1.

When input A is logic high, the output of IC1A is low and that of IC1B is high. The current then flows from the positive supply line to IC1A via T2, R9, and D8. T2 is switched on and D9 lights.

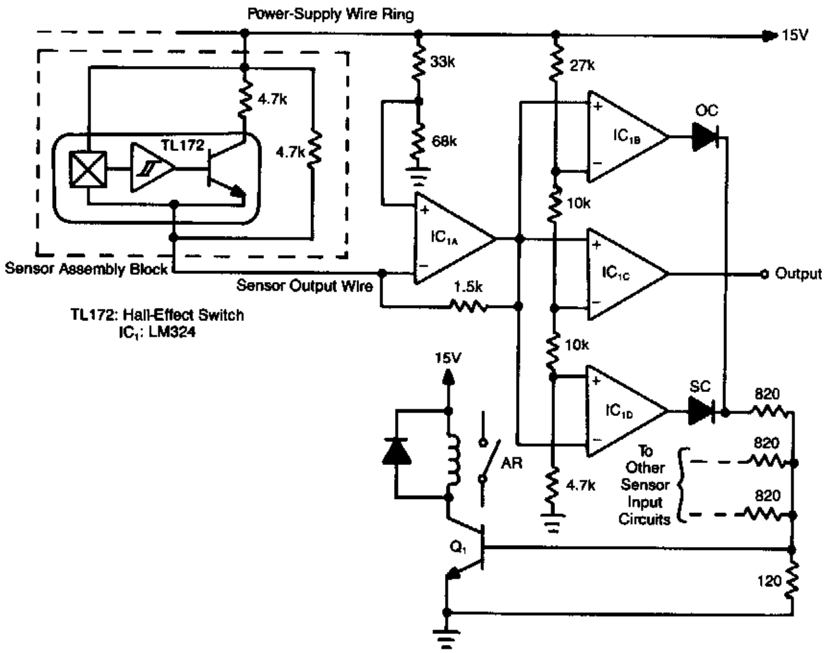
Because direct currents flow through R1/D1/R2 and R5/D3/R4, diodes D1 and D3 conduct and pass the VHF signal from input A to output D. At the same time, a direct current flows through R6/D4 so that D4 conducts. This arrangement ensures that any VHF signal at input C cannot reach the output via the parasitic capacitances of the relay contacts and the wiring.

When A is logic low, and IC1B is therefore low, the current flows from the positive supply line to IC1B via T1, R7, and D17. T1 is then switched on and D10 lights. At the same time, the two series-connected relays, Re1 and Re2, are energized, their contacts close, and the VHF signal at input C is fed to output D. Moreover, a direct current flows through R3/D2 so that D2 conducts. Any signal at input B is then shorted to ground via D2.

All resistors should be carbon-film types because these have a higher parasitic series inductance than metal-film resistors. Thus, the attenuation of the VHF signal caused by them is reduced to a minimum.

The attenuation losses caused by the diode junctions (5-10 dB) are somewhat larger than those caused by the relays. It is thus advisable to connect the antenna that provides the weaker signal (normally the domestic one) to input C. If the domestic antenna is equipped with an antenna amplifier, it can be supplied via terminal E. Diodes D5 and D6 protect the circuit against high-voltage spikes that occur during the on and off switching. The selector draws a current of approximately 65 mA.

($N+1$) WIRES CONNECT N HALL-EFFECT SWITCHES



EDN

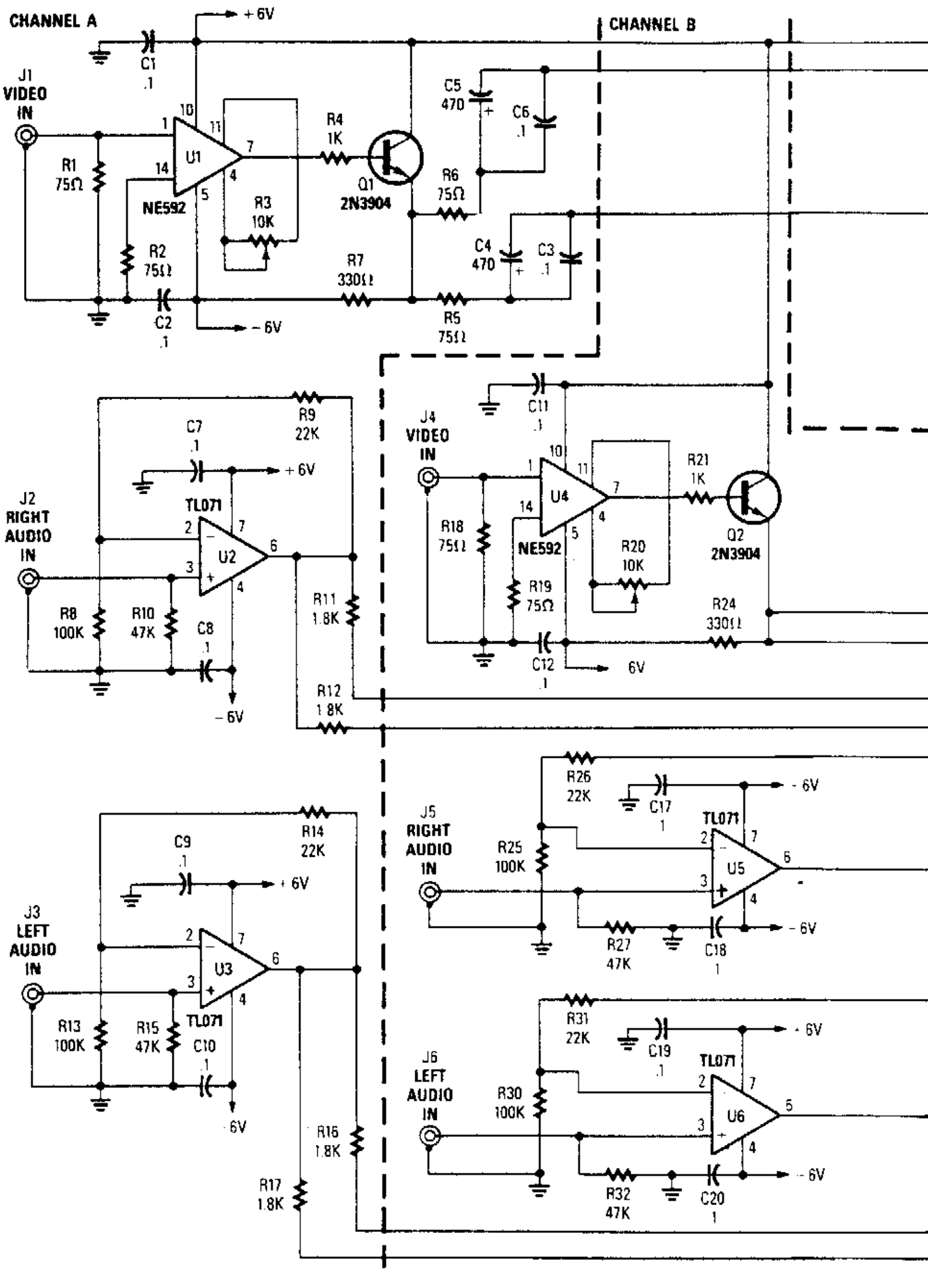
Fig. 84-2

Hall-effect switches have several advantages over mechanically and optically coupled switches. Their major drawback is that they require three wires per device. This circuit, however, reduces this wire count to $N+1$ wires for N devices.

Amplifier IC1A is configured as a current-to-voltage converter. It senses the sensor assembly's output current. When the Hall-effect switch is actuated, the sensor's output current increases to twice its quiescent value. Amplifier IC1B, configured as a comparator, detects this increase. The comparator's output goes low when the Hall-effect switch turns on.

The circuit also contains a fault-detection function. If any sensor output wire is open, its corresponding LED will turn on. If the power-supply line opens, several LEDs will turn on. A short circuit will also turn an LED on. Every time an LED turns on, Q1 turns on and the alarm relay is actuated.

AUDIO/VIDEO SWITCHER



POPULAR ELECTRONICS

This circuit is a two-channel baseband video switcher. Buffer amps U1/Q1, U4/Q4, and associated components produce a buffered 75-Ω video signal, which is routed to switching network K1/K2/K3. Relay K1 selects either of these two video amplifiers and feeds J7. K2 also routes the output of either video amplifier to K3, which passes the selected video channel to J8 or connects J9 to J8.

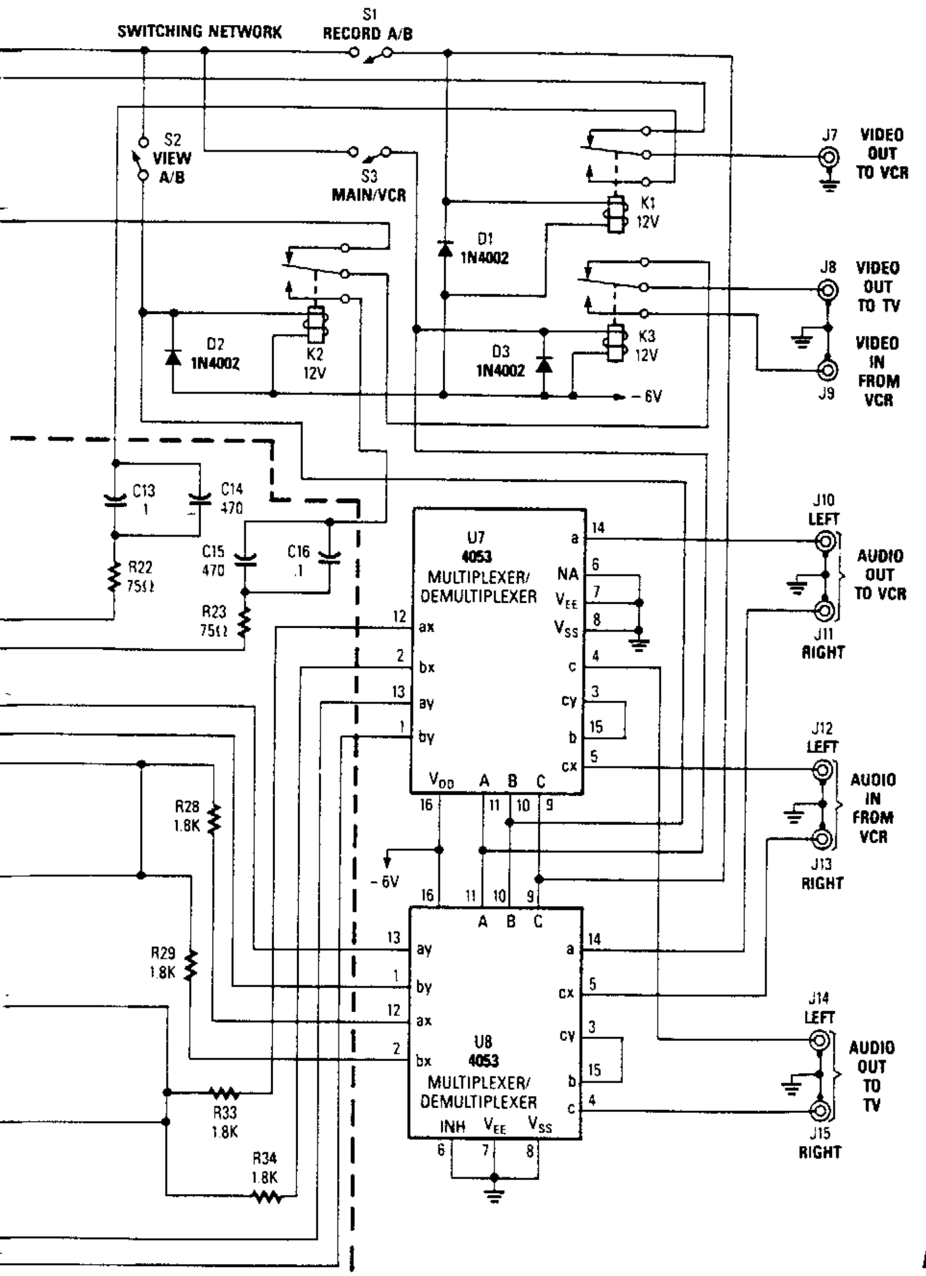
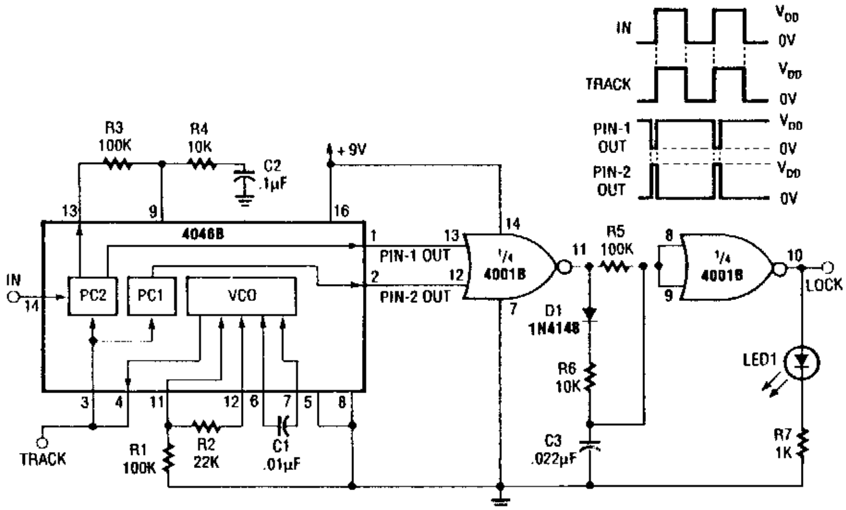


Fig. 84-3

U2 and J3 are audio amplifiers, which drive U7 and J8 (CD4053 analog switches) to route audio from J1 and J2 to either J14/J15, or J10/J11. Also, audio from J12/J13 can be routed to these jacks.

PRECISION NARROW-BAND TONE SWITCH



RADIO-ELECTRONICS

Fig. 84-4

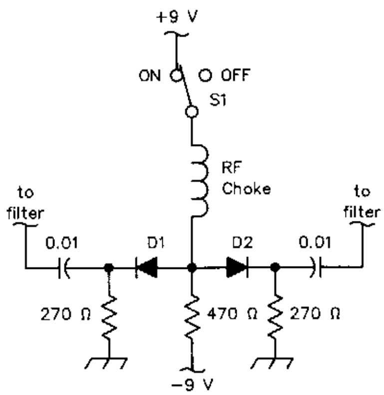
This signal tracker and lock detector combine to make a precision tone switch. Filter R3/R4/C2 determines signal capture and tracking range, as well as settling time.

Max. VCO frequency: R_1C_1

Min. VCO frequency: $(R_1+R_2)C_2$

Pin 9 voltage affects both. The minimum at pin 9 is 0 V and the maximum at pin 9 is V_{DD}. In the lock detector, the PC (phase comparator) outputs are pulses whose width is proportional to the phase difference between the two PC inputs. At lock up, the two PC outputs are almost mirror images. The output of IC1A remains low and IC1B is high. This lights LED1. If the loop is unlocked, the LED will not light.

DIODE RF SWITCH

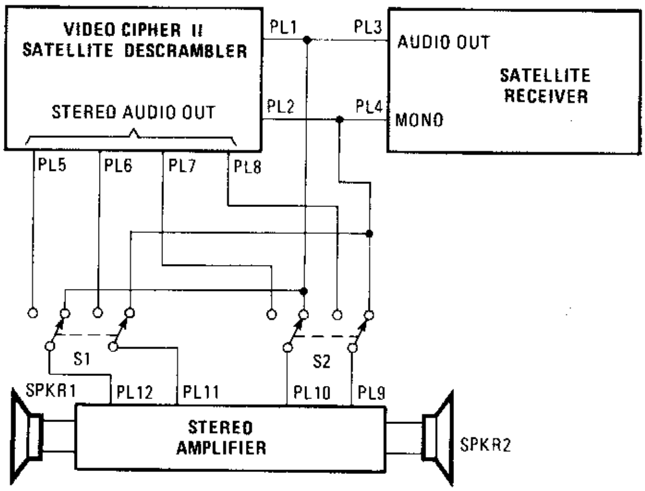


QST

Fig. 84-5

D1 and D2 can be IN914B- or HP2800-series diodes (for UHF). The loss is over 60 dB in the OFF state, and less than 3 dB at 3.5 to 30 MHz (using common IN914B diodes).

SATELLITE TV AUDIO SWITCHER

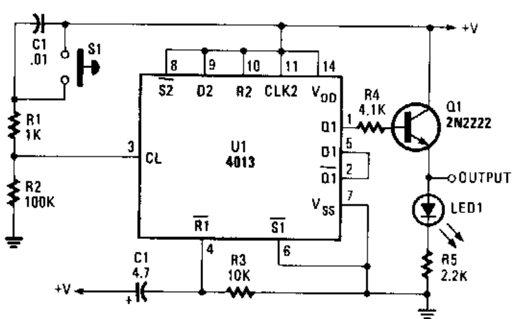


POPULAR ELECTRONICS **Fig. 84-6**

Some channels offer a separate audio mode (SAP) in a second language. It is usually transmitted on 6.8 MHz, which is the frequency used for unscrambled channels. The audio in the scrambled channels is transmitted along with the picture, so when the descrambler descrambles the signal, it also descrambles the audio in stereo. When the channel offers SAP, you'll find it on 6.8 MHz.

The switches are a pair of DPDT switches that have the toggle handles tied together. In one position, you hear the audio in stereo and in the other position, you hear the SAP. Just turn down the volume level on the TV and you can connect it to a stereo amplifier and a pair of speakers.

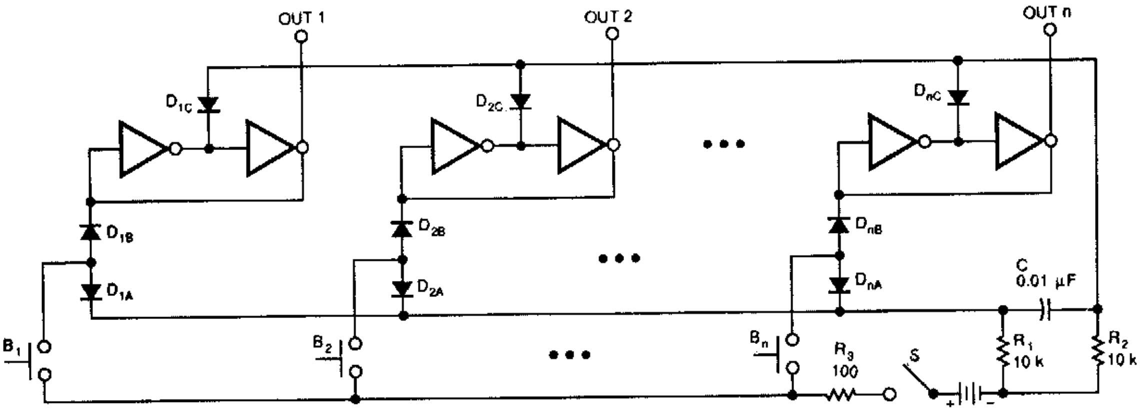
ON/OFF SWITCH



POPULAR ELECTRONICS **Fig. 84-7**

A CD 4013 dual-D flip-flop is used to drive an emitter-follower. This circuit can be used where a simple pushbutton on/off is desired.

SWITCHING CIRCUIT



EDN

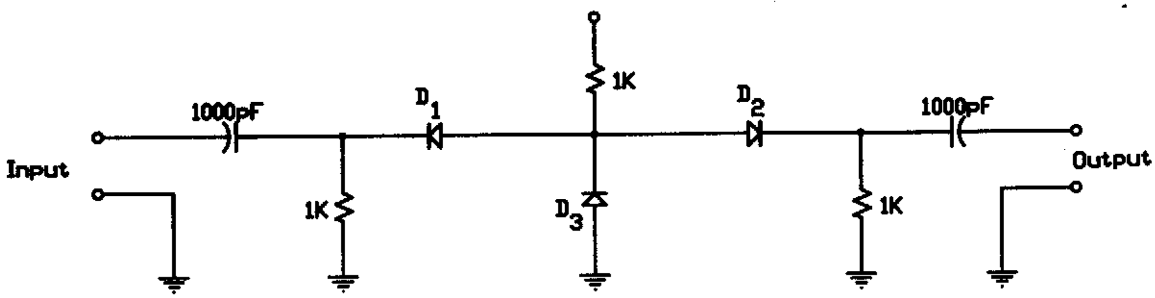
Fig. 84-8

This switching circuit acts like a bank of interlocked mechanical switches; pushing one of the buttons latches its corresponding output and unlatches a previously selected output. A pair of inverters forms a latch for each output.

Pressing button B1, for example, applies a positive pulse, via resistor diode D1B, to the input of the first output's, OUT1, latch. This positive pulse will set OUT1 high. Feedback locks OUT1's pair of converters in this HIGH state. Meanwhile, the pulse will also pass through diode D1A to the differentiator that is formed by C and R2. The differentiator will shorten the pulse.

The shortened pulse goes to all the latches and resets all of them, except the latch that sees the longer setting pulse. Obviously, if you press more than one button at once, more than one output will latch at once.

INEXPENSIVE VHF/UHF DIODE RF SWITCH



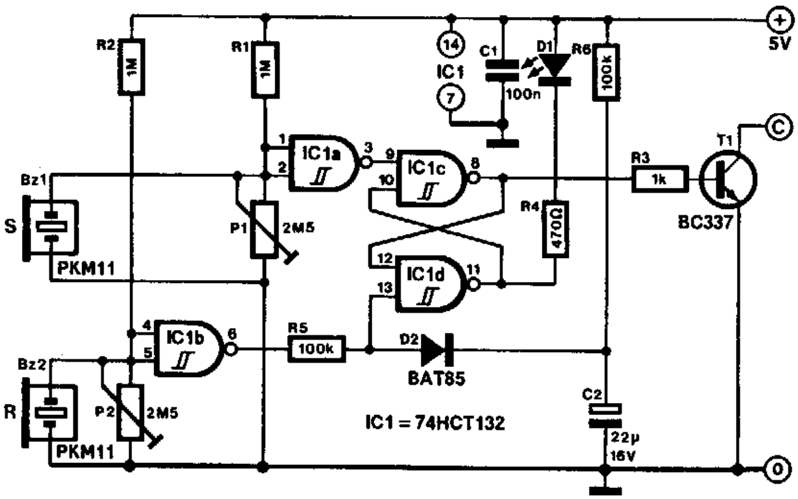
Diodes are 1N4148

RF DESIGN

Fig. 84-9

This circuit uses low-cost IN4148 diodes and exhibits about 1.5 dB insertion loss from 10 to 1 000 MHz with a few volts of negative bias. D3 conducts and D1/D2 are cut off, which results in 30 to 50 dB isolation. When a few volts of positive bias are applied, D1 and D2 are biased on and D3 is cut off. This circuit should be useful in applications where a low-cost RF switch is necessary.

MECHANICALLY CONTROLLED BISTABLE



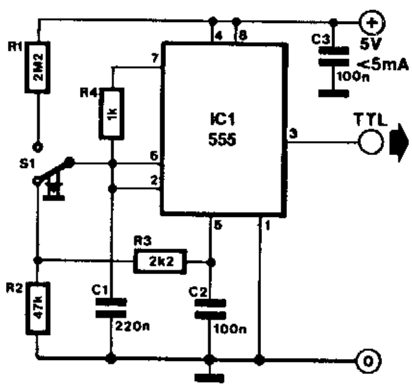
ELEKTOR ELECTRONICS *Fig. 84-10*

Applications for this mechanically set and reset bistable are found, among others, in antitheft devices and model railway crossings.

The transducers are formed by buzzer BZ1, which sets the bistable, and BZ2, which resets it. Their sensitivity is set with P1 and P2, respectively. The presets are adjusted correctly if the output of buffers IC1A and IC1B just toggles from high to low or vice versa.

If all have been set correctly, a slight tap of BZ1 will set the bistable. This tap causes T1 to switch on, which enables, for instance, a relay to be energized. At the same time, D1 lights. A tap on BZ2 or on its mounting resets the bistable, whereupon D1 goes out and T1 is switched off.

BOUNCE-FREE AUTO-REPEAT SWITCH



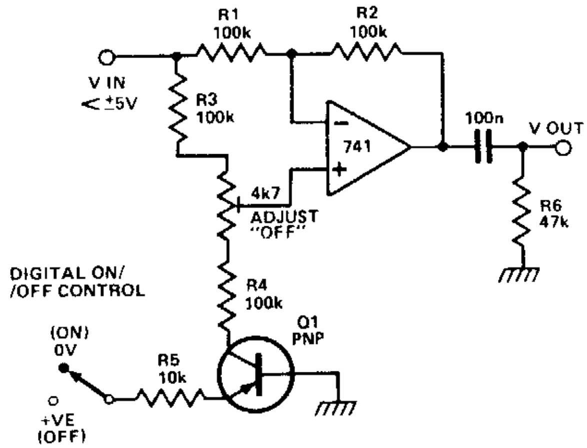
ELEKTOR ELECTRONICS *Fig. 84-11*

This switch that keeps pulsing as long as it is pressed is often required. The circuit here used the well known Type 555 for this purpose. Its output is a TTL-compatible signal.

Pin 5 of the timer has a potential of 67% of the supply voltage, U_{CC}. In the quiescent condition (switch not pressed), C1 charges via R2 and R3 to a voltage that is lower than that at pin 5, and thus is also lower than the toggle voltage.

When the switch is pressed, C1 is rapidly charged via R1 to the toggle voltage, upon which the timer emits a pulse. At the same time, the capacitor is discharged again via R4. As long as the switch is pressed, the circuit functions as an astable toggle and produces pulses. When it is released, the capacitor cannot charge to the toggle voltage.

TRANSISTOR TURNS OP AMP ON OR OFF



ELECTRONICS TODAY INTERNATIONAL *Fig. 84-12*

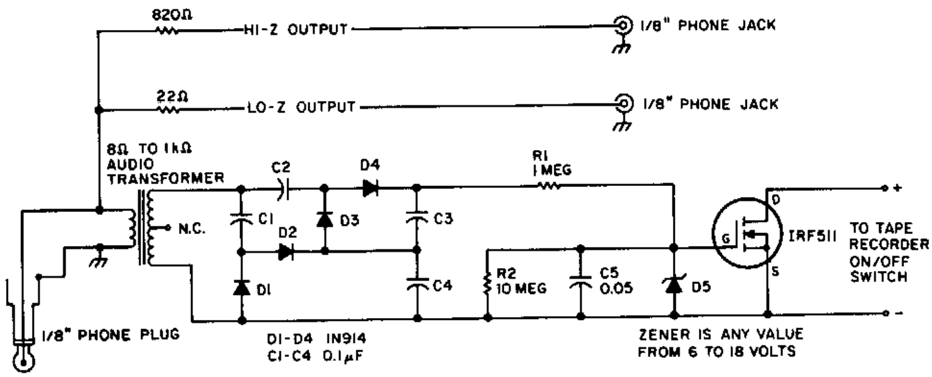
When transistor Q1 is switched off, the circuit behaves as a voltage follower. By applying a positive voltage to the emitter of Q1 via a 10 kΩ resistor, the transistor is made to turn on and go into saturation. Thus, the lower end of R4 is connected to ground. The circuit has not changed into that of a differential amplifier, except that the voltage difference is always 0 V. As long as the resistor ratios in the two branches around the op amp are in the same ratio, the output should be zero. A 47-kΩ resistor is used to null out any ratio errors so that the OFF attenuation is more than 60 dB. The high common-mode rejection ratio of a 741 enables this large attenuation to be obtained.

85

Tape Recorder Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 678. The figure number in the box of each circuit correlates to the entry in the Sources section.

AUDIO-POWERED TAPE RECORDER CONTROLLER

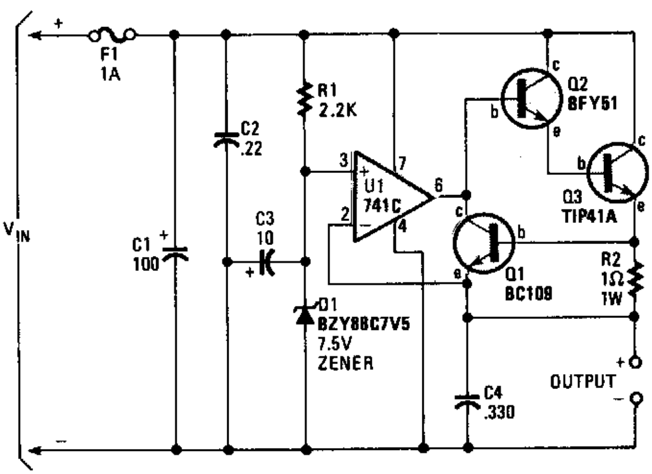


73 AMATEUR RADIO

Fig. 85-1

A tape recorder can be controlled by rectifying the audio input and driving an IRF511 power MOSFET to switch a tape recorder on when audio is present. This circuit was used with a communications receiver to record intermittent transmissions (such as aircraft, repeater output, etc.).

12-V AUTO-POWERED CIRCUIT FOR CASSETTE RECORDERS



HANDS-ON ELECTRONICS

Fig. 85-2

A regulator allows you to power a 7.5-V cassette recorder or other device from a 12-Vdc auto system. About 600 mA is available from the circuit. Q3 should be heatsinked because it dissipates up to 4 W. F1 should be a slow-blow fuse so that the surge caused by C1 does not cause unnecessary fuse failures.

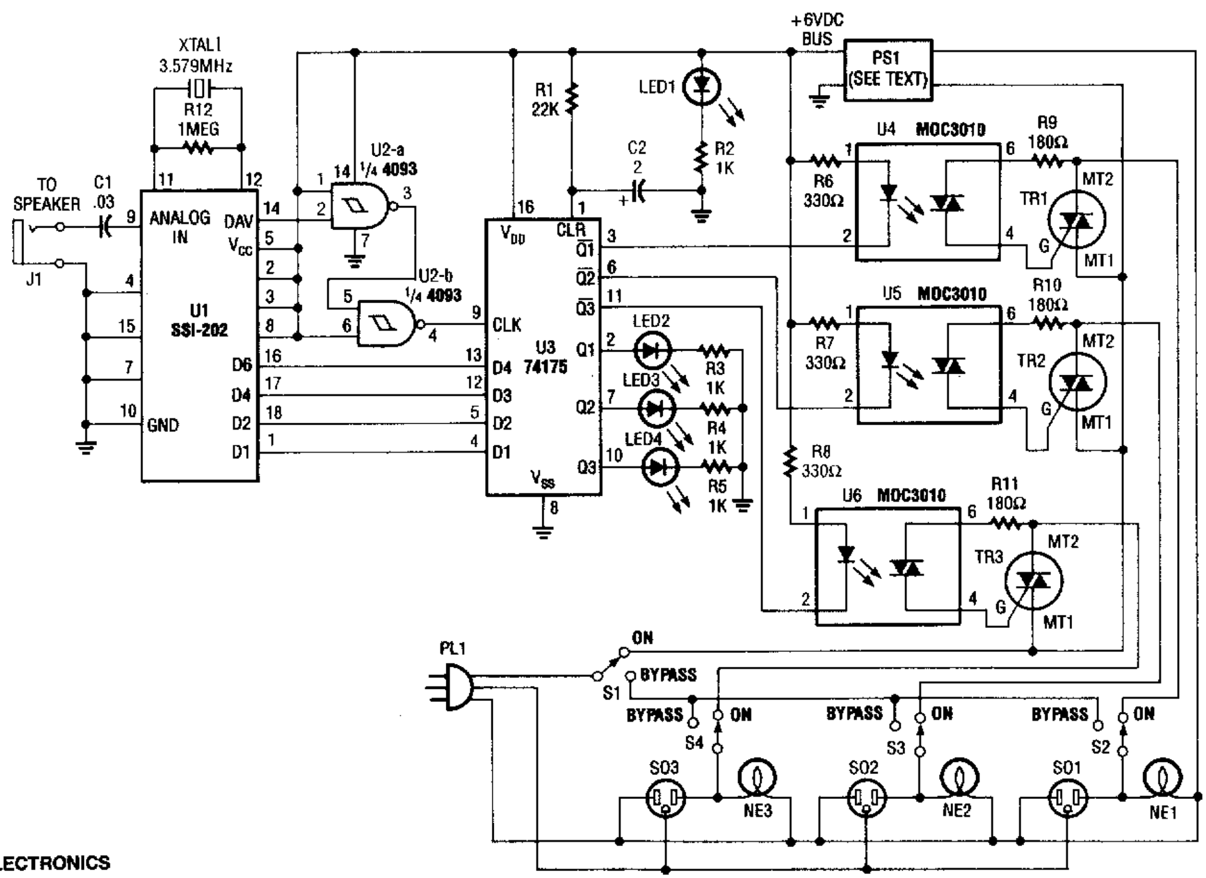
86

Telephone-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 678. The figure number in the box of each circuit correlates to the entry in the Sources section.

Telephone-Operated ac Power Switch
Telephone Toll Totalizer
Remote-Controlled Telephone/Fax Machine Switch
Telemonitor for Recording Phone Calls
Duplex Audio Link
Low-Power Touch-Tone Decoder
Telephone Speaker Amplifier
Telephone Ringer
Phone Message Flasher
Telephone Silencer
Telephone Intercom
Telephone Auto Record
Tell-A-Bell
Answering Machine Beeper
Telephone Visual Ring Indicator
Phone-In-Use Indicator
Telephone Amplifier
Extension Phone Ringer
Telephone Visual Ring Indicator
Call Tone Generator
Remote Ringer
Telephone Message Taker
Telephone Line-In-Use Indicator
Simple Ring Detector

TELEPHONE-OPERATED ac POWER SWITCH

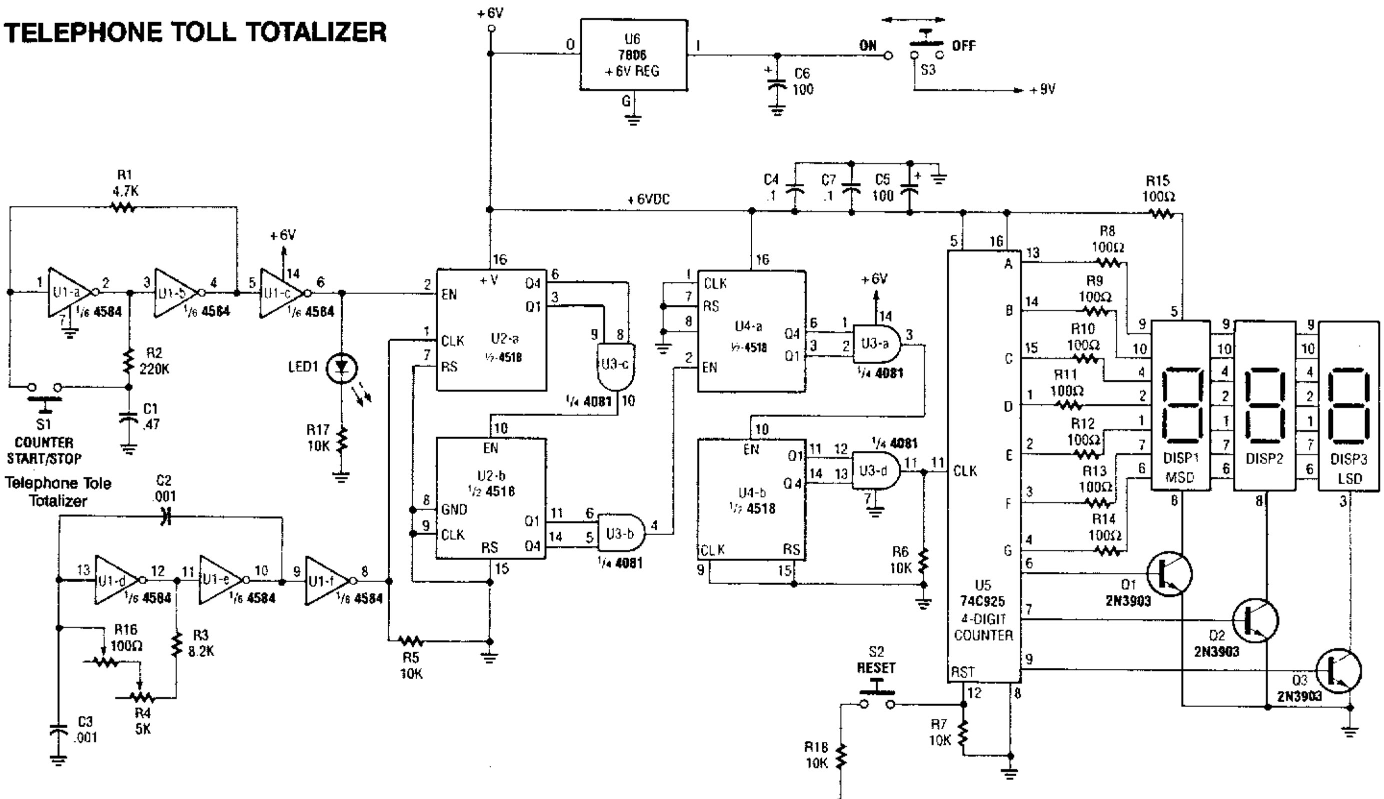


POPULAR ELECTRONICS

Fig. 86-1

Tones (from the DTMF) on the telephone line are detected by U1. When a valid tone is received, pin 14 (DAV) of U1 produces a positive pulse that is used to drive NAND gates U2A and U2B and then to latch binary data from U1 into Quad-D flip-flop U3 (U1 could be decoded into 16 bits, if required). The *Q* outputs of U3 drive optoisolators that control triacs TR1, TR2, and TR3. PS1 is a 6-V 150-mA dc adapter that operates from 120 Vac.

TELEPHONE TOLL TOTALIZER



POPULAR ELECTRONICS

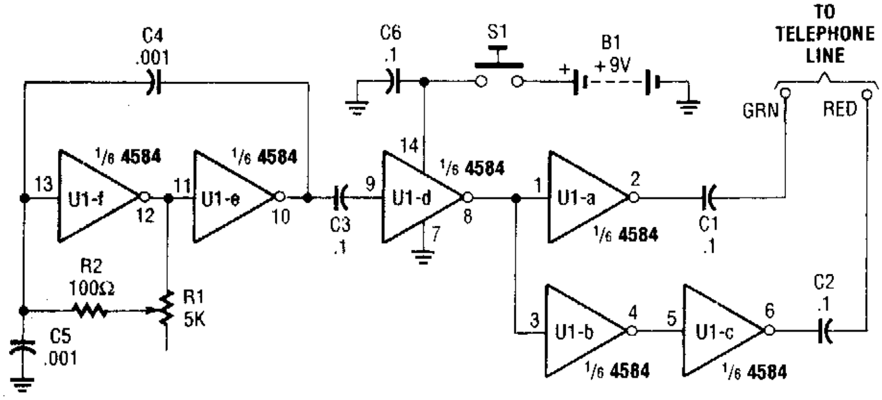
Fig. 86-2

The Telephone Toll Totalizer—built around two 4 518 dual synchronous up counters, a 74C925 4-digit counter, a 4 584 hex inverting buffer, and a 4 081 quad 2-input AND gate—is fairly simple.

Approximate toll charges can be calculated with this counter. It is started when dialing and stopped (manually) on hang-up. It is actually a counter that measures the time you are on the telephone. By calibrating it to the average cost/second of calls (get this from calculations you have done on your monthly phone bill), you can closely estimate your phone bill.

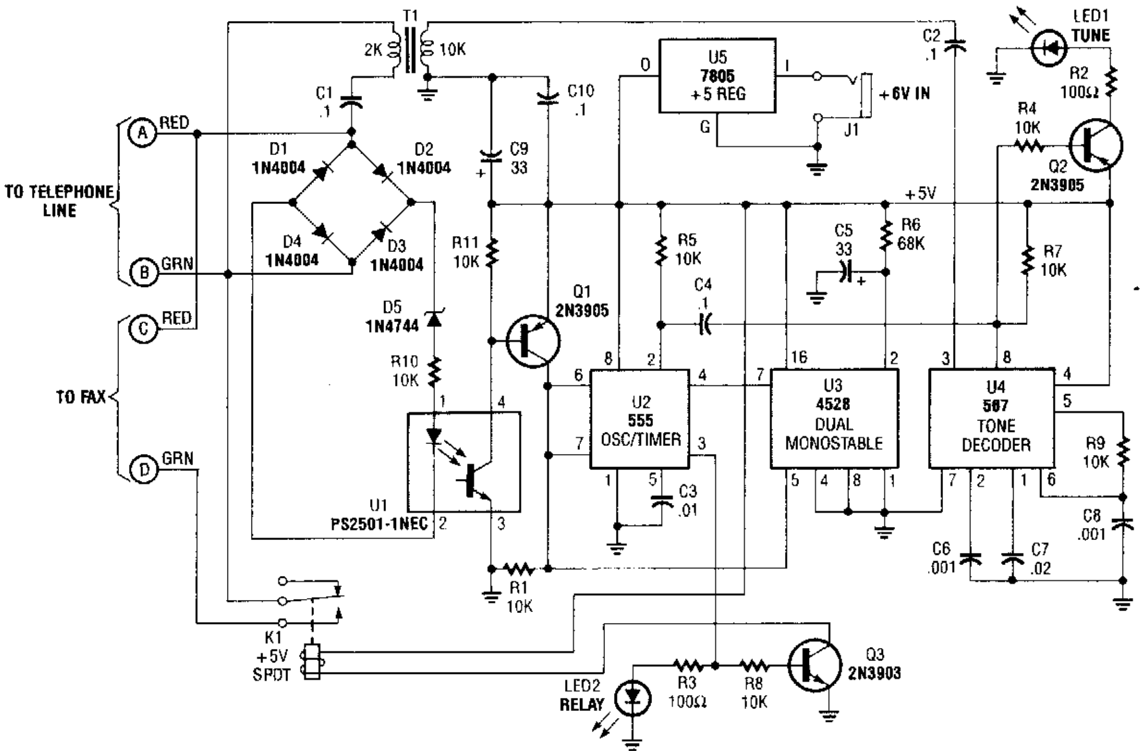
The circuit consists of an oscillator running at the 100 000 × frequency into the main counter (74C925). Typically, cost of telephone calls is 15 to 25 cents/minute so that the clock frequency (U1) is in the 25- to 40-kHz range. U2 and U4 with gates U3 form a ÷100 000 counter. The approximate cost in dollars and cents is read out on the multiplexed display, DISP 1, 2, 3. S2 resets the counter to zero after each use.

REMOTE-CONTROLLED TELEPHONE/FAX MACHINE SWITCH



POPULAR ELECTRONICS

Fig. 86-3(a)



POPULAR ELECTRONICS

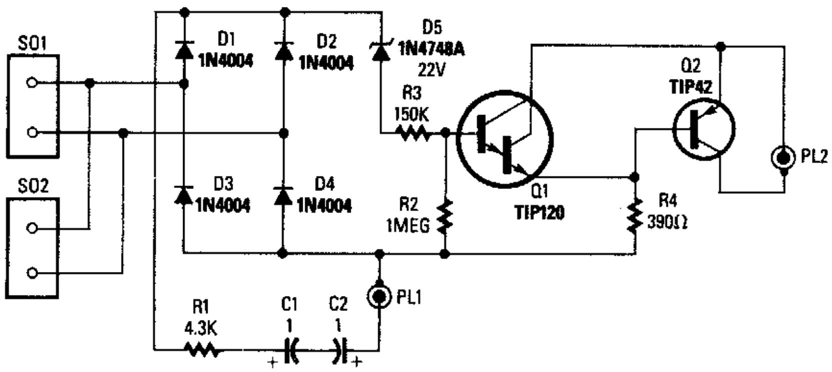
Fig. 86-3(b)

This system uses a transmitter at around 100 kHz (see Fig. 86-3(a)) to control a remote receiver. A line splitter can be used to connect the transmitter to the telephone line in use. The transmitter is a CMOS oscillator and has output buffer stages to drive the telephone line.

When the receiver (Fig. 86-3(b)) detects the off-hook condition (the line voltage drops from about 48 V to less than 10 V). Optocoupler U1 has the LED extinguished. This enables timer U2. When the transmitter is activated, tone decoder U4 detects the 100-kHz signal and outputs a low signal, which lights indicator LED1. LED1 is also used to set the transmitter frequency. Also, U2 is triggered. U2 is configured for the latching condition. U2 feeds the base of Q3, turns it on and energizes relay K1, which switches in the fax machine to the telephone line.

U3 prevents U2 from being accidentally triggered by transients on the telephone line. When the phone is lifted off-hook and it resets U2 after about one or two seconds of delay, transients are allowed to subside before U2 is reset, and it waits for a negative pulse on pin 2 to turn on. When U2 turns on, Q3 is biased on, which activates change-over relay K1.

TELEMONITOR FOR RECORDING PHONE CALLS

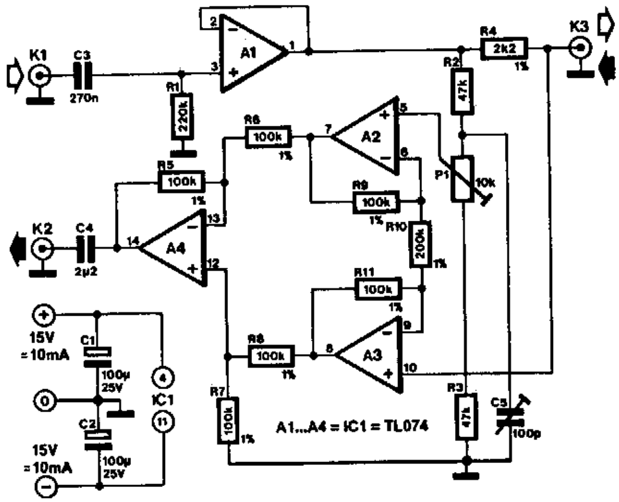


POPULAR ELECTRONICS *Fig. 86-4*

This circuit switches a tape recorder via PL2. When on-hook, D5 conducts, turns on Q1, and cuts off Q2. Ringing voltage will also cause D5 to conduct; C2 and C1 should be rated 150 V or higher. When phone is off-hook, the 10 V or so present on the line will not break down D5, and therefore Q1 is off and Q2 is biased on. PL2 connects to the remote control jack on the tape recorder. Audio is taken from PL1.

Caution: Use either a battery tape recorder or an FCC/CSA/UL-approved ac adapter-powered tape recorder. This precaution is to avoid inadvertent 120 Vac on the telephone line.

DUPLEX AUDIO LINK



ELEKTOR ELECTRONICS *Fig. 86-5*

Duplex communication is, of course, not a new technique: it has been used, for instance, in telephone systems for many years. Those systems, however, use transformers to achieve duplex—this circuit does it with the aid of electronics.

The principle is fairly simple. Two senders impose signals (U_1 and U_2, respectively) on to the audio cable. The voltage across the cable is then $(U_1 + U_2)/2$. The receivers at both sides of the cable deduct their side's sender signal from the cable signal: the result is that the signal is sent from the other end of the cable. This principle is the basis of the circuit shown. Notice that a similar circuit is required at either end of the link.

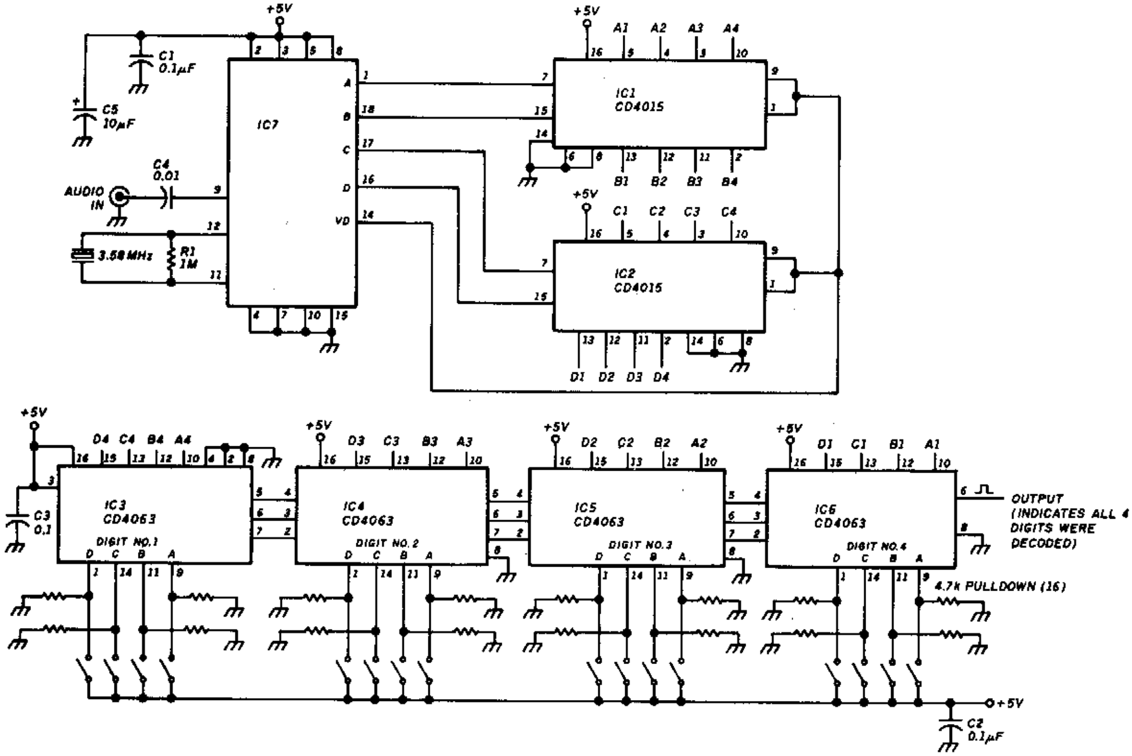
Op amp A1 is connected as a buffer amplifier and serves as sender. The send signal is imposed on the cable via R4. Terminating the cable by R4 results in the voltage across the cable being only half the voltage output of A1. This does not detract from the operation of the circuit, however. At the same time, R4 ensures that signals emanating from the other end of the link cannot get to the output of A1; if they could, they would be short-circuited by the output.

The receiver is a differential amplifier consisting of op amps A2 through A4. The quality of the differential amplifier depends largely on the resistors used with the op amps; 1% types are, therefore, essential.

The cable signal, $(U_1 + U_2)/2$, is applied to one input of the differential amplifier and the (halved) output signal of A1 to the other. Because the differential amplifier has a gain of 6 dB, the received signal applied to K2 has the same level as the original signal.

The circuit is calibrated by connecting the cable to it and to its twin circuit, then injecting a 1-kHz sinusoidal signal and a 5-Vrms level to its input. The input bus of the other circuit must be short-circuited during the calibration. Adjust P2 for minimum signal at K2. Next, increase the frequency of the input signal to 10 kHz and adjust C5 for a minimum signal at K2. Repeat the procedure with the other circuit. The signal suppression at 1 kHz is of the order of 80 dB; at 20 kHz, it is approximately 60 dB.

LOW-POWER TOUCH-TONE DECODER

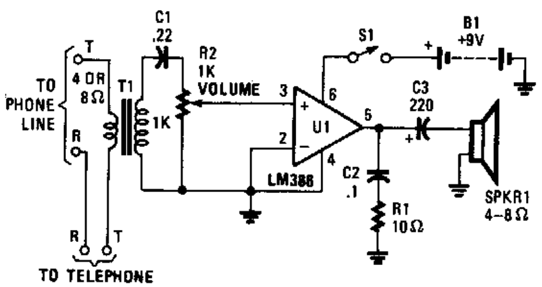


HAM RADIO

Fig. 86-6

This decoder will respond to a preselected 4-digit DTMF number. IC7 is a Radio Shack IC device (part #276-1303). The logic is all CMOS. The digits are selected by SW1 and SW2, a pair of 8-position DIP switches.

TELEPHONE SPEAKER AMPLIFIER

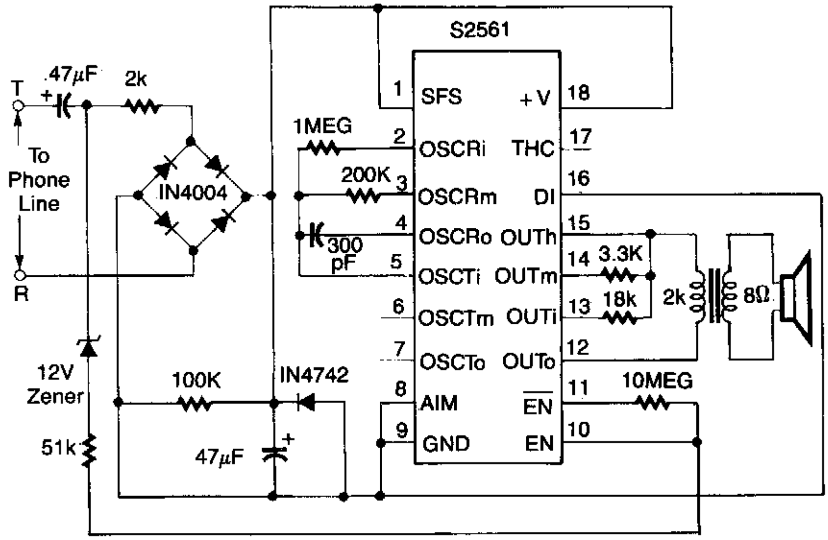


This simple telephone amplifier (which can be switched off for privacy) allows everyone in the room to listen to your telephone conversations.

POPULAR ELECTRONICS

Fig. 86-7

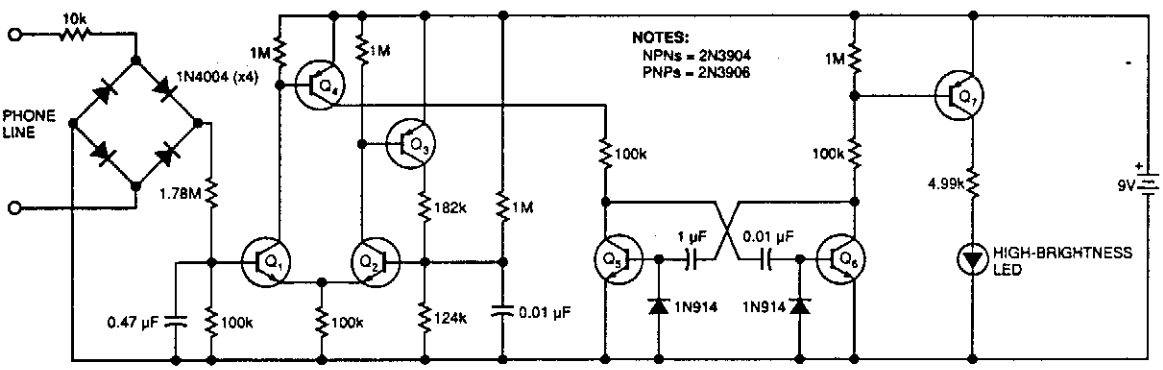
TELEPHONE RINGER



RADIO-ELECTRONICS *Fig. 86-8*

Using an AMI P/N S2561 IC, the circuit shown can be either powered by a battery or the telephone line in use. Output is about 50 mW.

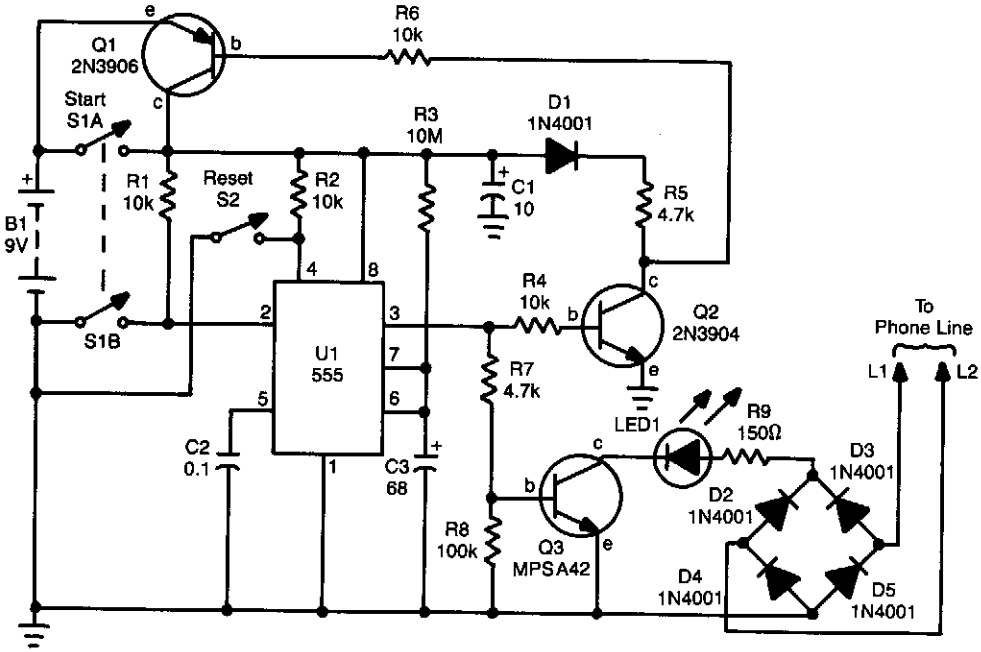
PHONE MESSAGE FLASHER



EDN *Fig. 86-9*

This circuit flashes an LED to indicate that your phone rang during your absence. A differential amplifier with hysteresis (Q1, Q2, and Q3) detects high line voltage (ringing), which turns on Q4, multivibrator Q5/Q6, and flashes the LED via Q7. Q1 and Q2 remain on until the phone-line voltage drops to less than 9 V, which indicates an off-hook condition.

TELEPHONE SILENCER

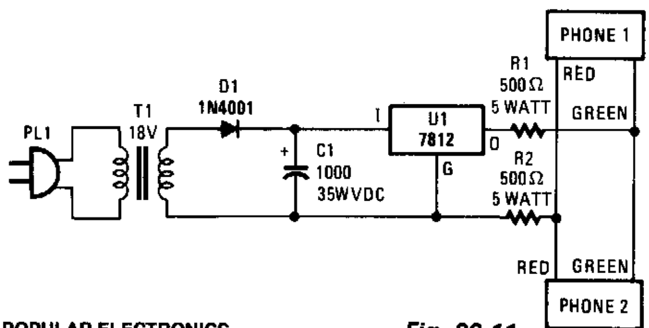


WELS' THINK TANK

Fig. 86-10

If you are busy and cannot answer or do not wish to answer your phone, this circuit will give a busy signal without you having to leave the phone off the hook. After a predetermined time, the circuit is deactivated. U1 forms an astable multivibrator that can be set for a time up to 10 minutes by values of R3 and C3. When S1A is depressed, U1 starts, and Q1 latches, which powers the circuit. At the end of a time interval determined by R3 and C3, Q2 and Q3 cut off and remove power from the circuit. During the operation, S3 throws a 150-Ω resistor across the phone line, which simulates an off-hook condition.

TELEPHONE INTERCOM

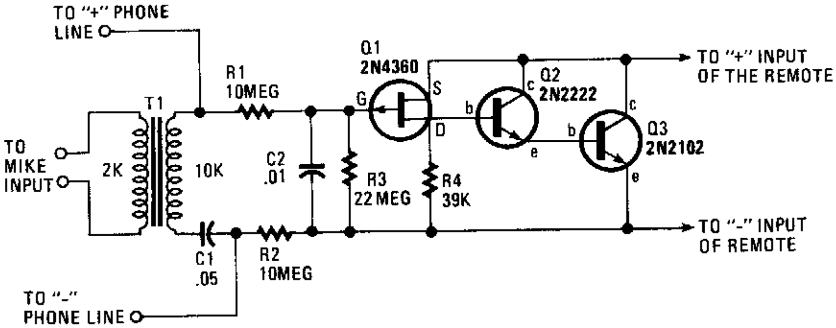


Two telephones can be used as an intercom setup with this simple power-supply arrangement. The 500-Ω resistors maintain line balance.

POPULAR ELECTRONICS

Fig. 86-11

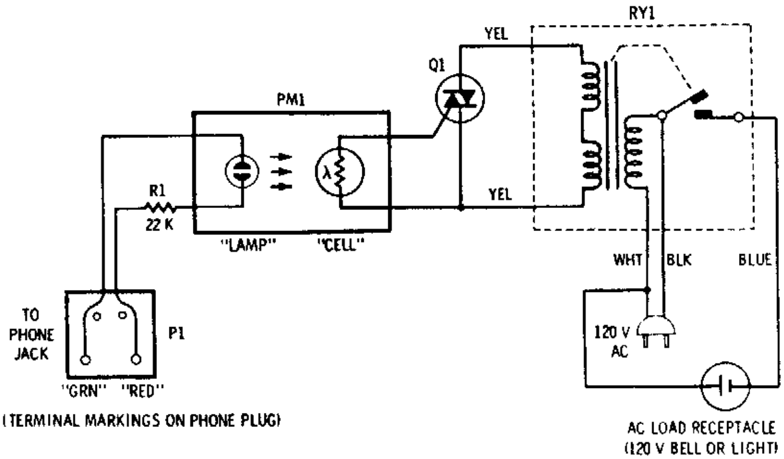
TELEPHONE AUTO RECORD



HANDS-ON ELECTRONICS *Fig. 86-12*

The circuit requires neither a battery nor an ac supply to make it work. Set the recorder to the record position and when the telephone is taken off the hook, the recorder starts to record everything. When the phone is on the hook, the voltage across the phone lines is about 48 Vdc. When it is taken off the hook, the line voltage drops below 10 V. When the line voltage is near 48 V, the FET is biased off and no current can flow through Q2 and Q3. When the receiver is off the hook, the voltage drops, and allows Q1 to conduct. This action turns on Q2, Q3, and the cassette recorder.

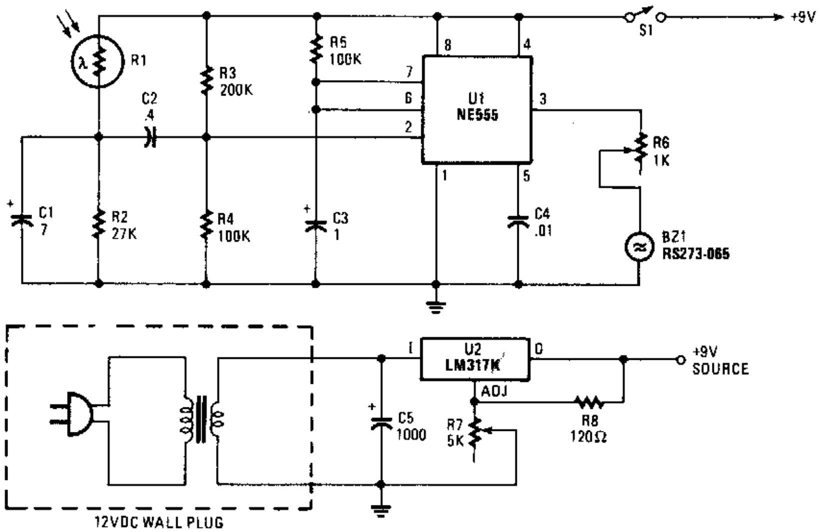
TELL-A-BELL



TAB BOOKS *Fig. 86-13*

This accessory connects to phone and will activate an attention-getting 120-V bell whenever a ringing voltage appears on the phone line. The four-terminal transducer has a neon bulb close to a photocell, which are both enclosed in a light-tight tube. When the lamp is off, the cell is dark and its resistance is very high. When a ringing voltage appears on the phone line, the neon bulb glows brightly and illuminates the photocell whose resistance then drops to about 1000 Ω. The photocell is in the gate circuit of a triac that will turn on whenever the cell resistance drops. The triac is connected across the switching terminals of the isolation relay. Thus, the relay closes and applies 120 V to the output socket whenever the triac is on.

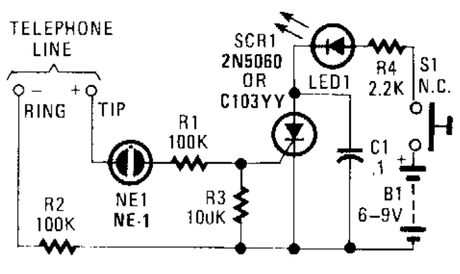
ANSWERING MACHINE BEEPER



POPULAR ELECTRONICS *Fig. 86-14*

When the light on the answering machine blinks, the resistance of photoresistor R1 charges, triggers the timer (U1), and generates 0.2-s pulse that activates BZ1. R1 is optically coupled to the LED on the answering machine and is properly light shielded.

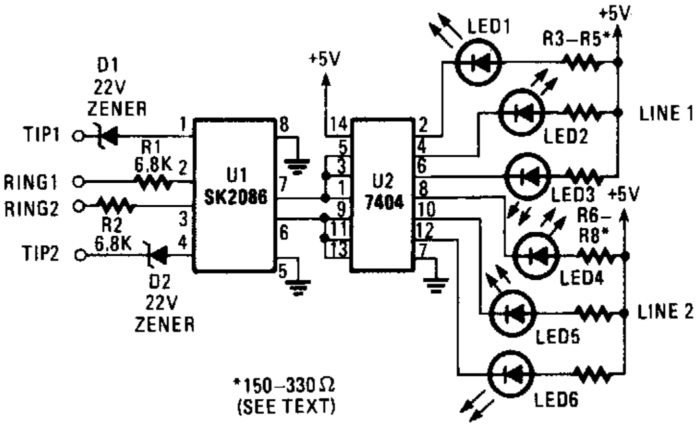
TELEPHONE VISUAL RING INDICATOR



POPULAR ELECTRONICS *Fig. 86-15*

This circuit will indicate the receipt of a call. When the telephone rings, a 100- to 120-V ring signal breaks over NE1, and causes SCR1 to trigger. This causes LED1 to light until the SCR1 is turned off by depressing S1.

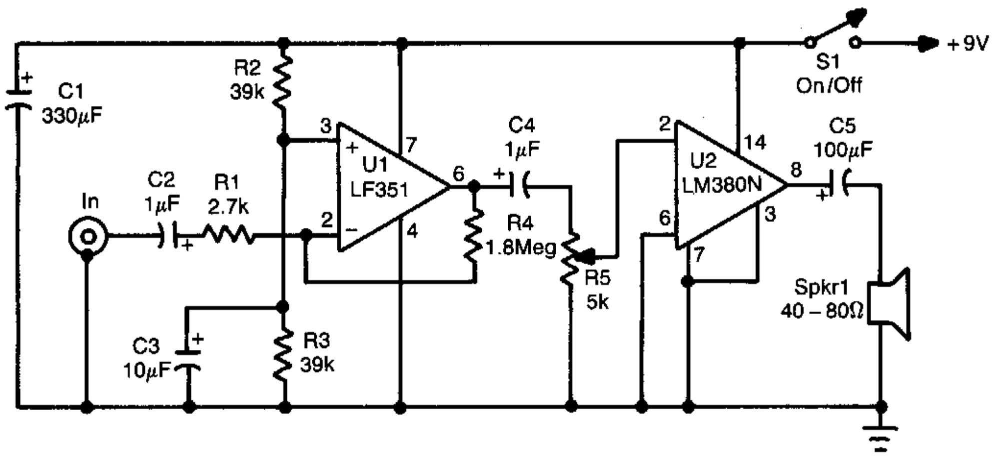
PHONE-IN-USE INDICATOR



POPULAR ELECTRONICS *Fig. 86-16*

The circuit receives its power from a 15-V wall adapter (not shown). The circuit takes advantage of the fact that the phone line voltage drops from 48 to 10 V when an extension is taken off the hook. When the voltage on a line drops, the optoisolator/coupler is turned off so that the inputs to the line-1 hexinverters (U2 pins 1, 3, and 5) float high. The corresponding outputs (U2 pins 2, 4, and 6) go low and light the line-1 LEDs.

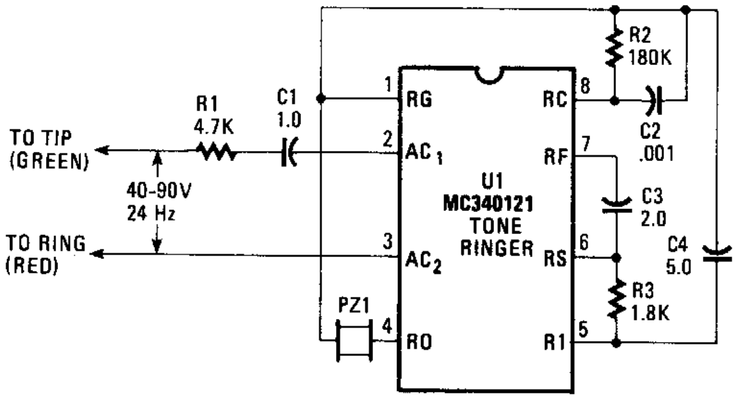
TELEPHONE AMPLIFIER



POPULAR ELECTRONICS *Fig. 86-17*

This amplifier can be used in telephone work or where a simple speech amplifier is required. The frequency response can be varied by the value of C_2, C_4, and addition of a capacitor across R4 (≈ 33 pF for voice band) to limit the HF response.

EXTENSION PHONE RINGER

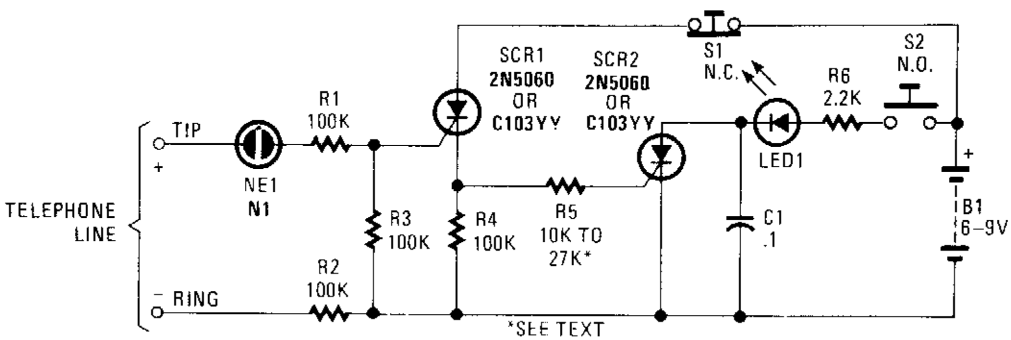


HANDS-ON ELECTRONICS

Fig. 86-18

The ac ringing voltage (typically 40 to 90 V at 26 Hz) is rectified by U1, the tone-ringer IC, and is used to drive that IC's internal tone-generator circuitry. The tone-generator IC includes a relaxation oscillator (with a base frequency of 500, 1 000, or 2 000 Hz) and frequency dividers that produce the high- and low-frequency tones, as well as the tone-warble frequency. An on-board amplifier feeds a 20-Vpp signal to the transducer.

TELEPHONE VISUAL RING INDICATOR

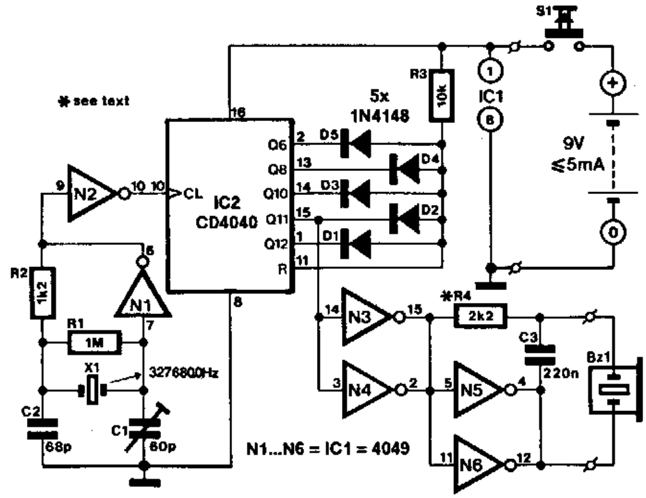


POPULAR ELECTRONICS

Fig. 86-19

In this circuit, the ringing voltage on a telephone line causes NE-1 to break over, triggering SCR1, which in turn triggers SCR2. If a call has been received, depressing S2 will cause LED1 to light. Depressing S1 resets the circuit. This circuit has the advantage of lower battery drain because LED1 is not left on continuously after a ring signal, but only when S2 is depressed.

CALL TONE GENERATOR



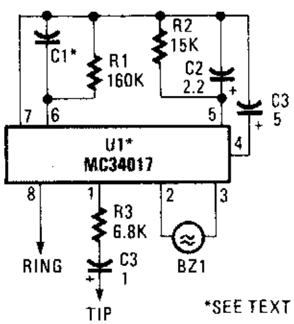
ELEKTOR ELECTRONICS *Fig. 86-20*

Amateur VHF relay stations are normally actuated by a 1 750-Hz call tone. This might give problems if the relevant sending equipment has no internal call-tone generator, if it does not have sufficiently accurate frequencies, or if the tone duration is not long enough to securely energize the relevant relay.

These problems can be overcome by the stand-alone generator described here. Simply placed in front of the microphone, it makes absolutely certain that the relay station is actuated. The generator consists of a quartz oscillator, a frequency counter and a buffer-amplifier—all contained in just two CMOS ICs. It is powered by a 9-V (p-p) battery, from which it draws a current of around 5 mA.

Gates N1 and N2 form an oscillator that is controlled by a 3.276 80-MHz crystal and provides clock pulses to IC2, which is connected as a programmable scaler. Diodes D1 through D5 determine the divide factor of 1 872. Counter output Q1 provides the wanted 1 750-Hz signal, which is buffered by N3 through N6 before being applied to a piezoelectric buzzer. Capacitor C3 suppresses any harmonics, while R4 determines the volume of the output signal.

REMOTE RINGER



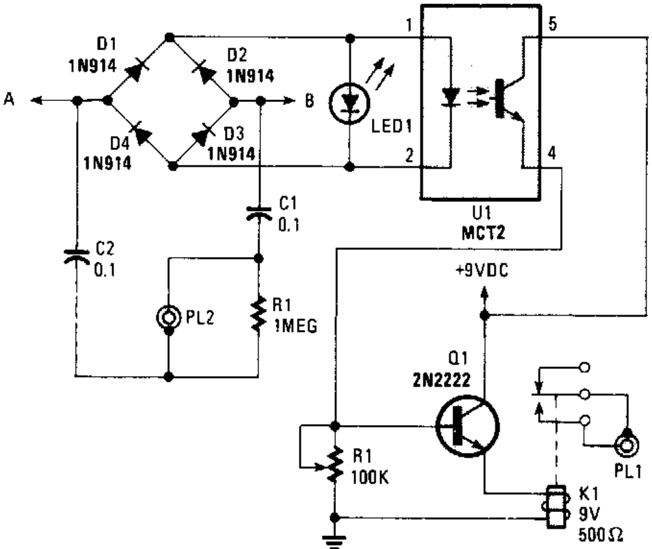
POPULAR ELECTRONICS *Fig. 86-21*

A telephone bell circuit using a Motorola MC34017 can be built from a few components. C1 and U1 depend on the type of ring required.

U1	C1	Ring
MC34017-1	1 000 pF	1 kHz
MC34017-2	500 pF	2 kHz
MC34017-3	2 000 pF	500 Hz

Select the version of MC34017 and C1 from this table.

TELEPHONE MESSAGE TAKER

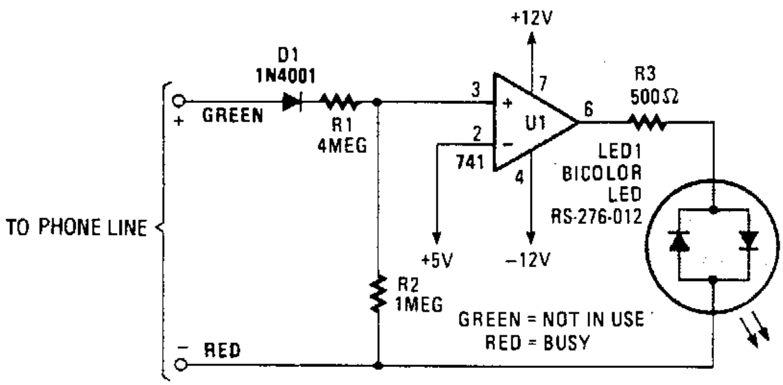


POPULAR ELECTRONICS

Fig. 86-22

This circuit operates on the ringing voltage of the telephone to trigger a tape recorder to record messages. K1 can be made to latch using extra contacts if the tape recorder requires a constant-contact closure.

TELEPHONE LINE-IN-USE INDICATOR

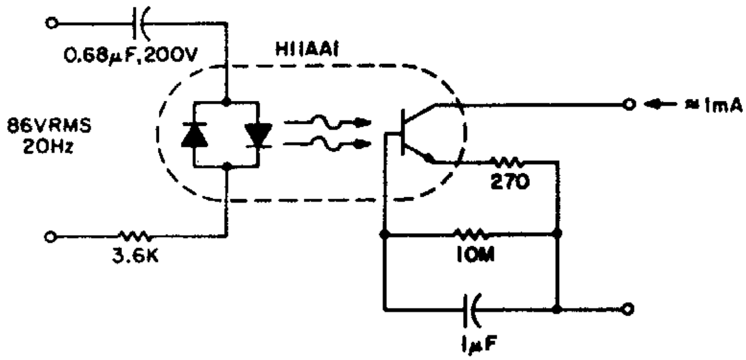


POPULAR ELECTRONICS

Fig. 86-23

When a telephone line is not in use, about 48 V appears across the line, which drops to about 10 V or less when the line is in use. This circuit switches a bicolor LED as an indicator.

SIMPLE RING DETECTOR



GE

Fig. 86-24

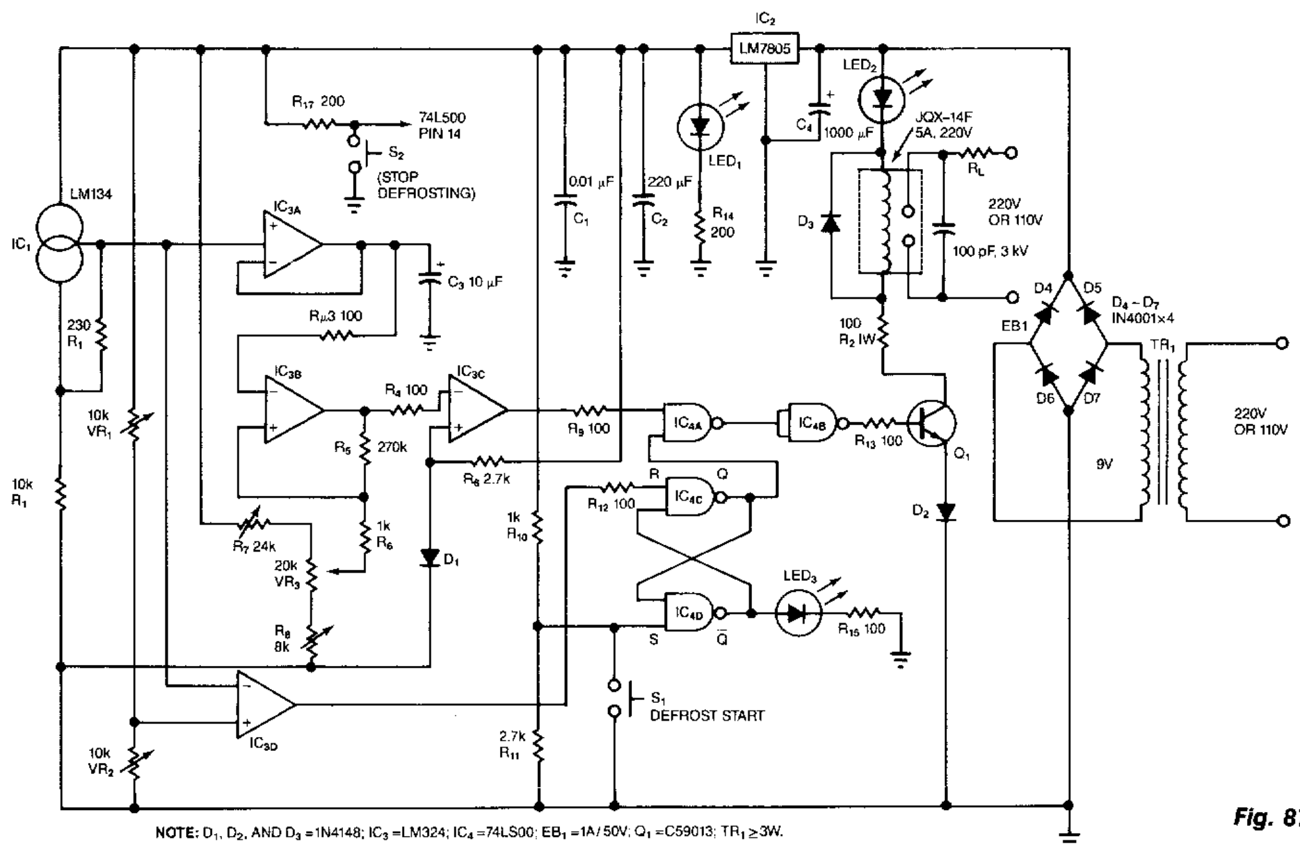
This circuit detects the 20-Hz approximately 86-Vrms ring signal on telephone lines and initiates action in an electrically isolated circuit. Typical applications include automatic answering equipment, interconnect/interface and key systems. The detector is the simplest and provides about a 1-mA signal for a 7-mA line, which loads for 0.1 s after the start of the ring signal. The time-delay capacitor provides a degree of dial-tap and click suppression, and filters out the zero crossing of the 20-Hz wave.

87

Temperature Controls

The sources of the following circuits are contained in the Sources section, which begins on page 679. The figure number in the box of each circuit correlates to the entry in the Sources section.

TEMPERATURE CONTROLLER WITH DEFROST CYCLE



EDN

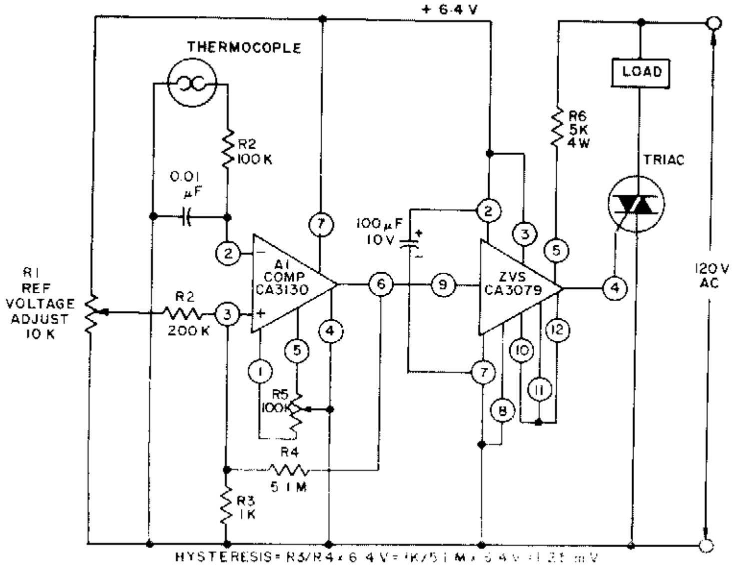
NOTE: D_1, D_2, AND D_3 = 1N4148; IC_3 = LM324; IC_4 = 74LS00; EB_1 = 1A/50V; Q_1 = C59013; $TR_1 \geq 3W$.

Fig. 87-1

This temperature controller has a range of −50 to +150°C and permits defrosting. R7, VR3, and R8 set the controller's trip point. S1 initiates defrosting, S2 cancels defrosting. VR1 and VR2 set the defrost-temperature trip point.

The LM134, IC1, is a thermal sensor. One section of IC3, a quad op amp, buffers the sensor's output. The other section functions as Schmitt trigger and buffer for the normal-cycle circuitry and as a comparator for the defrost-cycle circuitry. If you wish to control a heater rather than a refrigerator, omit the final inverter in Q1's base circuit. You must select LM7805s that have outputs between 4.95 and 5.05 V, or the circuit might not work.

THERMOCOUPLE TEMPERATURE CONTROL

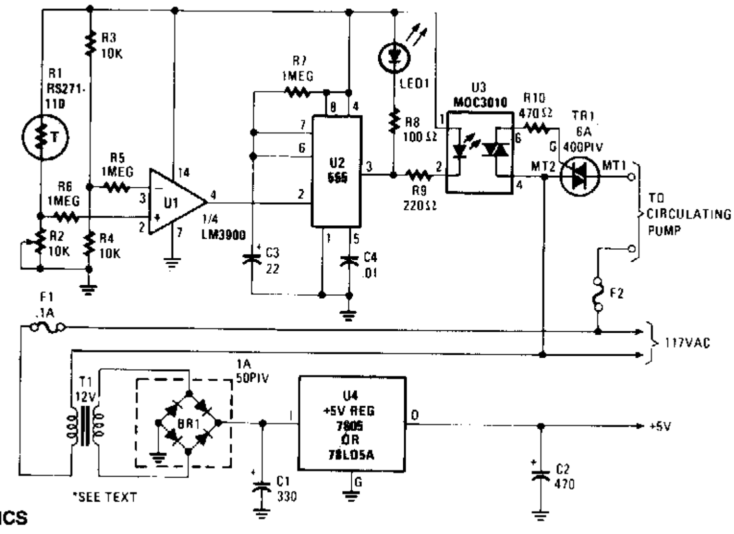


GE/RCA

Fig. 87-2

This control, with zero-voltage load switching, uses a CA3130 BiMOS op amp and a CA3079 zero-voltage switch. The CA3130, used as a comparator, is ideal because it can "compare" the low voltages generated by the thermocouple to the adjustable reference voltage over the range of 0 to 20 mV.

TEMPERATURE CONTROLLER



POPULAR ELECTRONICS

Fig. 87-3

A thermistor (R1) is compared with a reference (R2) in a Wheatstone-bridge circuit. Comparator U1's output goes high, which triggers U2. U2 is a delay of about 25 s. After 15 s, LED1 lights, U3 actuates, triac TR1 triggers, and turns on a hot water pump. This system was used with a hot-water heater.

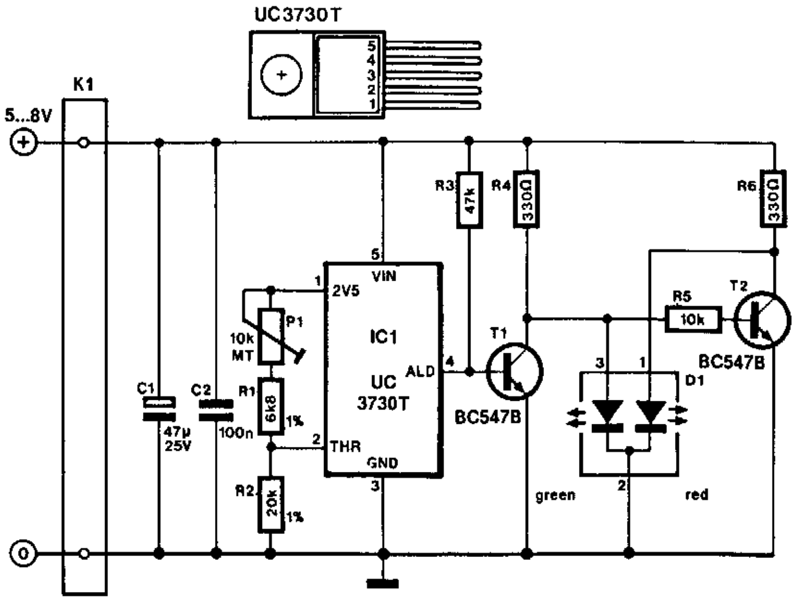
88

Temperature Sensors

The sources of the following circuits are contained in the Sources section, which begins on page 679. The figure number in the box of each circuit correlates to the entry in the Sources section.

Thermal Monitor
Simple Temperature Indicator
Under-Temperature Switch
Temperature Sensor
Over-Temperature Switch
Transistor Sensor Temperature Measurer

THERMAL MONITOR



ELEKTOR ELECTRONICS

Fig. 88-1

Unitrode's UC1730 family of integrated circuits is designed for use in a number of thermal monitoring applications. Each IC combines a temperature transducer, a precision reference, and a temperature comparator to allow the device to respond with a logic output if temperatures exceed a predetermined level.

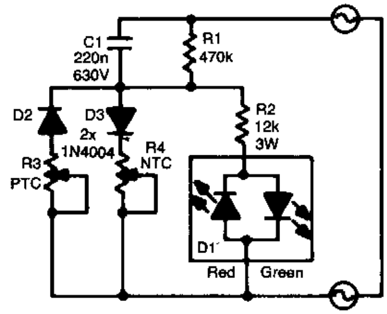
The monitor presented here is based on a UC3730T and it is intended to be fitted to a heatsink. Although the supply to the device can be as high as 40 V, 5- to 8-V is chosen here, because it is normally readily available in the equipment where the monitor will be used (power amplifiers, power supplies, etc.).

The threshold temperature, T_t, in °C, is determined by:

$$T_t = \frac{2.5R_2}{0.005(R_1+R_2+P_1)-273.15}$$

The temperature can be preset with $P1$ to values between −1°C and +100°C. The indicator is formed by a bicolor LED and controlled by transistors T1 and T2. Resistors R4 and R5 limit the current through the LED. When the temperature of the heatsink is below the threshold temperature, the ALD (alarm delay) output, pin 4, is logic low so that T1 is switched off and the green LED lights.

SIMPLE TEMPERATURE INDICATOR



ELEKTOR ELECTRONICS

Fig. 88-2

For the absolute measurement of temperatures, a thermometer is indispensable. However, in many situations, an absolute value is not needed and a relative indication is sufficient. It would be a further advantage if a green light would indicate that all is well as far as temperature is concerned. As the temperature rises, the light should change color slowly to indicate that the equipment is getting too hot.

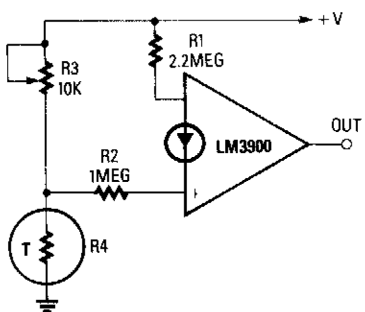
This circuit does this and works directly from the mains. The indicator proper is a two-color LED (D1), while the sensor is a combination of a negative-temperature coefficient (NTC) and a positive-temperature coefficient (PTC) resistor (R4 and R3, respectively).

At a relatively low temperature, the value of R_3 is low and that of R_4 is high. During the positive half cycle of the mains voltage, a voltage will exist across R3/D2 that is sufficiently high to cause the green section of D1 to light. The value of R_3 has been chosen to ensure that during the negative half cycle of the mains voltage, the potential across it is too low to cause the red section of D1 to light.

If the temperature rises, the value of R_4 diminishes and that of R_3 rises. Slowly, but surely, the green section will light with lesser and lesser brightness. At the same time, the red section lights with greater and greater brightness until ultimately only the red section will light.

Resistor R2 and capacitor C1 ensure that the current drawn by the LEDs does not become too large. This arrangement keeps the dissipation relatively low. Both R3 and R4 should be of reasonable dimensions—approximately 6 mm in diameter, not less. At 25°C, the NTC must be 22 to 25 kΩ and the PTC must be 25 to 33 Ω. The circuit should be treated with great care because it carries the full mains voltage.

UNDER-TEMPERATURE SWITCH

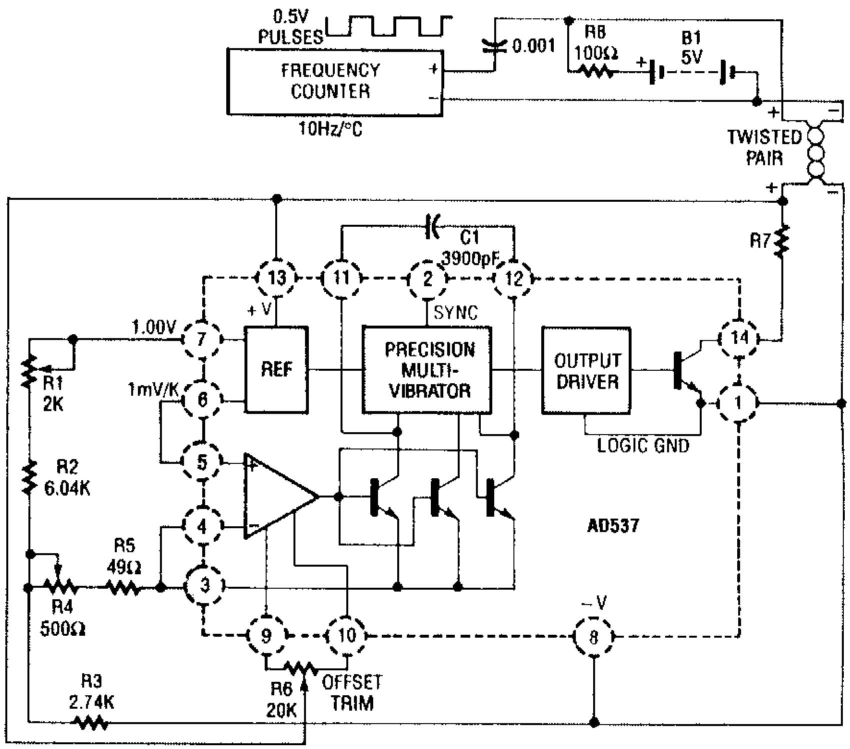


RADIO-ELECTRONICS

Fig. 88-3

The reference current is fed from the supply voltage via R1, to the inverting terminal, and the variable (noninverting) current is supplied from the junction of R3 and R4. Because the value of R_1 is approximately double that of R_2, and generates a current that is proportional to the supply voltage, the trip temperature (preset via R3) is independent of the supply voltage.

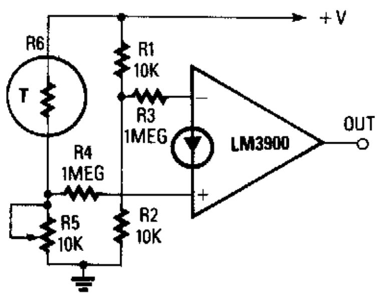
TEMPERATURE SENSOR



RADIO-ELECTRONICS *Fig. 88-4*

The AD537 uses its two reference outputs—one fixed at 1 V, the other of which varies with temperature (1 mV per °K). At 0°C, the 1-V reference multiplied by 0.273 will balance this voltage and produce a zero output. The scale in this circuit is 10 Hz/°C. Output from, as well as power to, the circuit, is fed via a two-wire twisted pair. A frequency counter is used as a readout.

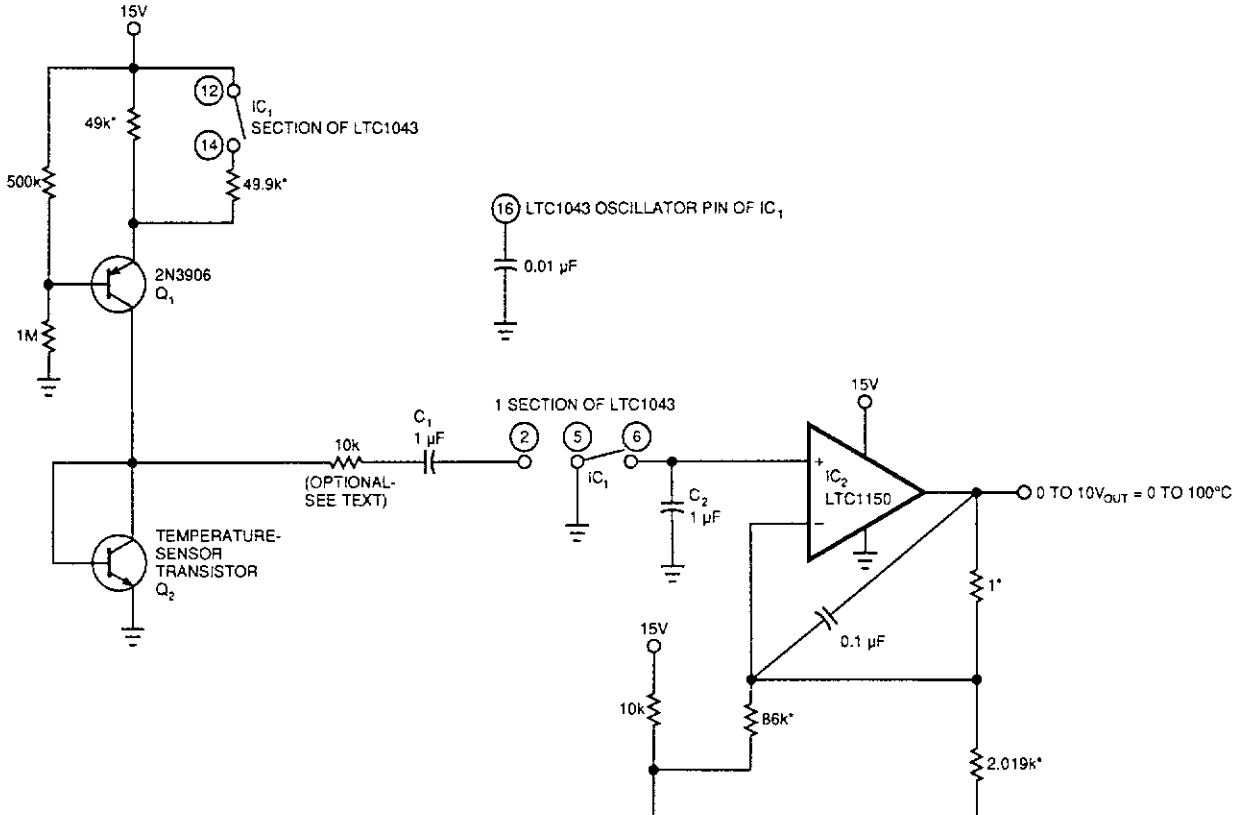
OVER-TEMPERATURE SWITCH



RADIO-ELECTRONICS *Fig. 88-5*

The output goes high when a preset temperature is exceeded. A fixed half-supply reference voltage feeds a reference current to the inverting input, and a variable current is fed to the noninverting input. Resistor R6 is a negative-temperature-coefficient (NTC) thermistor, so the potential at the junction of R5 and R6 rises with temperature. The op amp will switch high when that voltage exceeds the half-supply value. The trip temperature can be preset via R5.

TRANSISTOR SENSOR TEMPERATURE MEASURER



NOTES:
* = 0.1% FILM RESISTOR
SENSOR TRANSISTOR MAY BE ANY SMALL-SIGNAL NPN-2N2222, 3904, ETC.

EDN

Fig. 88-6

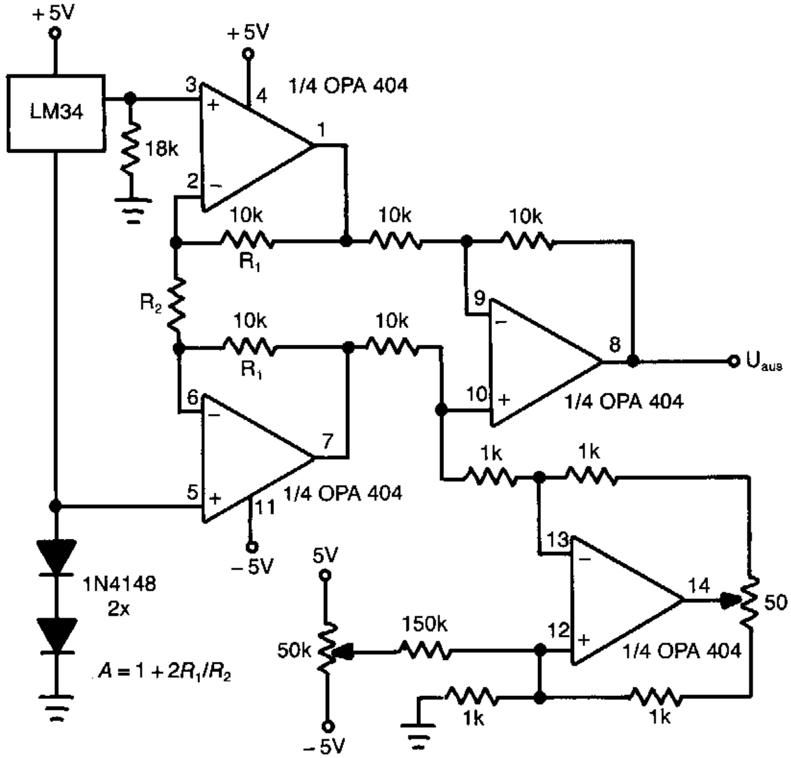
Using the fact that the V_{BE} of a transistor shifts 59.16 mV per decade of current at 25°C. This constant is 0.33%/°C. This V_{BE}-vs.-current relationship holds true regardless of the V_{BE} absolute value. IC, an LT1043, acts as an oscillator and switches a current source (Q1) at a 10:1 ratio. The stepped 10:1 current drive is translated to temperature by IC2, Q2, and the associated components. Accuracy is ±1%. No compensation is needed if Q2 is changed.

89

Thermometer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 679. The figure number in the box of each circuit correlates to the entry in the Sources section.

INEXPENSIVE LINEAR THERMOMETER



ELECTRONIC ENGINEERING

Fig. 89-1

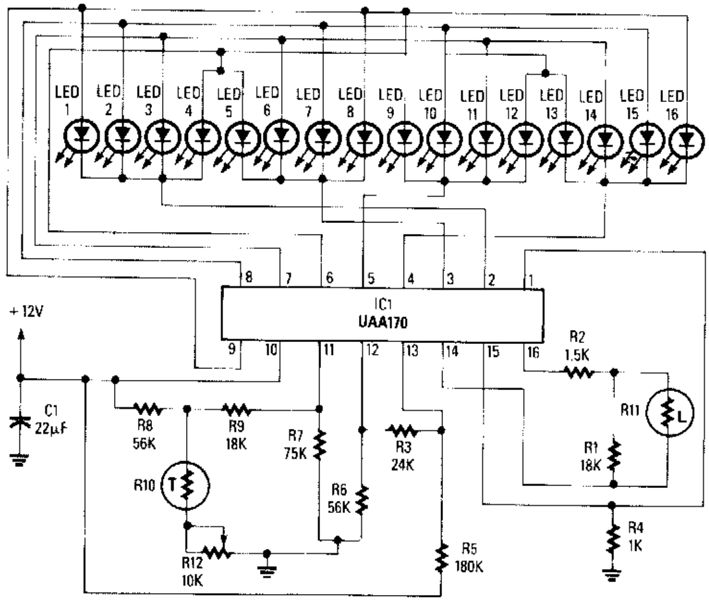
As a sensor in the design of this linear thermometer, the LM34 is used. The output is the difference between two base-emitter voltages ΔV_{EB} of two transistors operated at different collector-current densities. The voltage difference (ΔV_{EB}) is:

$$\Delta V_{EB} = V_{EB1} - V_{EB2}$$
$$= \frac{kT}{q} \ell n \, (I_{C1}/I_{C2})$$

where the current densities are I_{C1} and I_{C2}, k is Boltzmann's constant, and q is the electron charge. Because all factors, including the ratio I_{C1}/I_{C2} are constant, the output of the LM34 (National Semiconductor) is a linear function of T in the range over $-50°$ to 300°F, which provides the output voltage of 10 mV/°F with a max. nonlinearity of ± 0.35°F.

The output of the LM34 is amplified by a three-op-amp instrumentation amplifier. The fourth op amp gives the possibility to control the offset voltage of the amplifier. The gain A of the instrumentation amplifier can be set to any desired value by the choice of the resistance R_2 only.

ELECTRONIC THERMOMETER

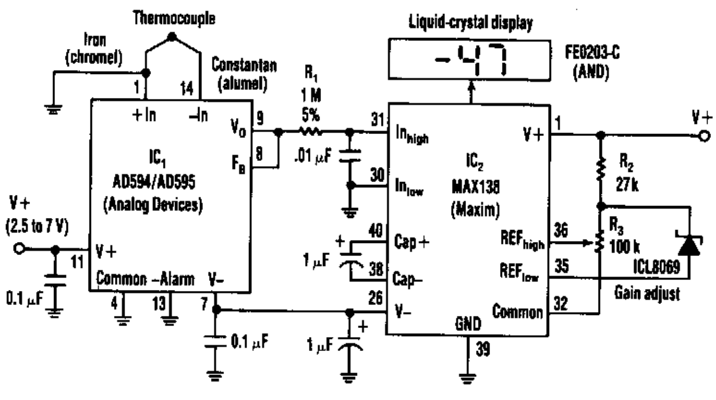


RADIO-ELECTRONICS

Fig. 89-2

This thermometer uses an NTC thermistor (R10) to produce a dc voltage that decreases with temperature, to drive an IC that lights one of the 16 LEDs as a function of this voltage. R11 is a light-dependent resistor that adjusts the LED brightness as a function of ambient light.

SINGLE-dc SUPPLY THERMOMETER

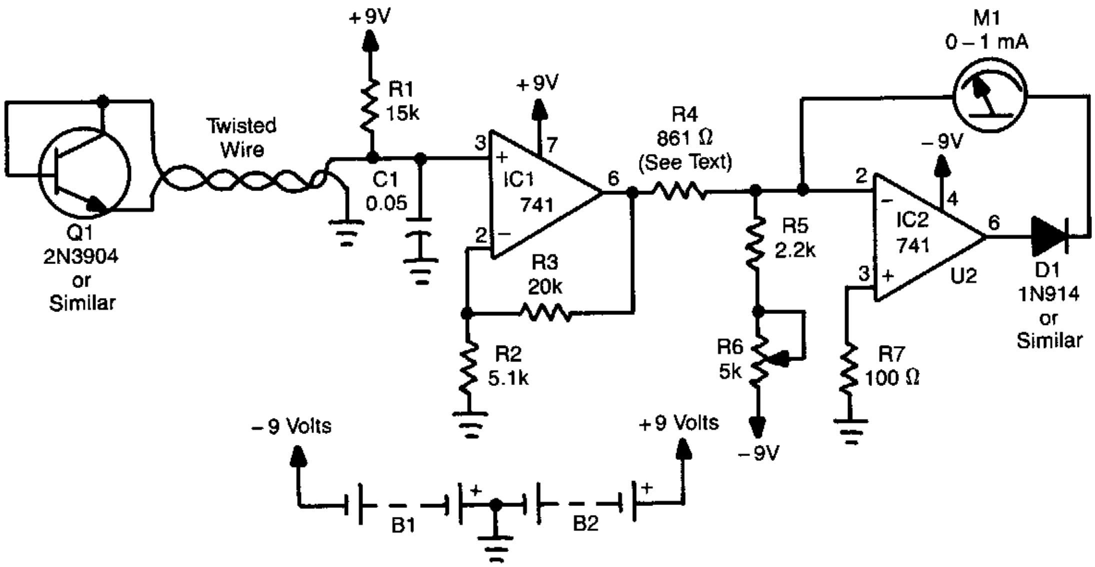


ELECTRONIC DESIGN

Fig. 89-3

Using a J-type thermocouple, this circuit can indicate temperatures from −350° to 400° with a 6-V supply or −50 to +100° with a 3-V lithium battery. The AD954 produces 10 mV/°C output to the MAX 138 digital voltmeter chip, which drives the LCD display.

ELECTRONIC THERMOMETER



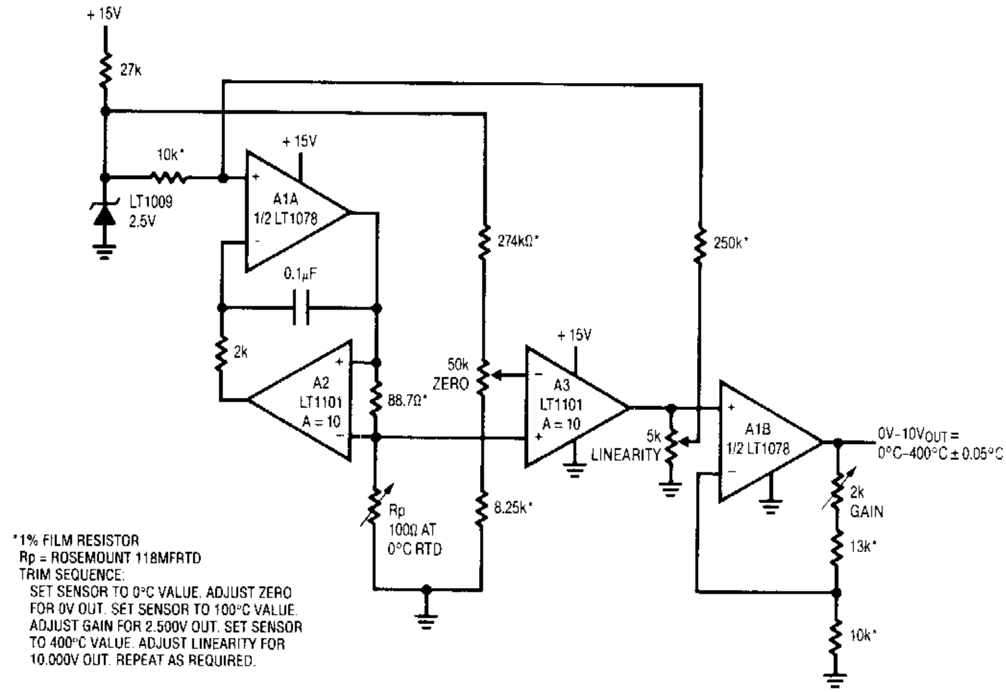
GERNSBACK PUBLICATIONS

Fig. 89-4

This thermometer is capable of measuring temperatures from −30 to +120°F. A diode-connected 2N3904 transistor forms a voltage divider with R1. The transistor is used as the temperature sensor and, for best results, it should be connected to the rest of the circuit with twisted wire, as shown. As temperature increases, the voltage drop across the transistor changes by approximately −1.166 mV/°F. As a result, the current at pin 3 of IC1, a 741 op amp with a gain of 5, decreases as the temperature measured by the sensor increases. A second 741 op amp, IC2, is configured as an inverting amplifier.

Resistors R5 and R6 are used to calibrate the current. At a temperature of about −30°F, the current through R4 (formed by connecting a 910- and a 1 600-Ω resistor in parallel) should equal the current through R5 and R6. A temperature of −30°F will result in a meter reading of 0 mA, while a temperature of 120°F will result in a meter reading of 1 mA. Divide the scale between those points into equal segments and mark the divisions with the appropriate corresponding temperatures. If you divide it into 150 equal segments, for instance, each division will equal one degree. Calibration is completed by placing the sensor in an environment with a known temperature, such as in an ice-point bath. The freezing point of water is approximately 32°F. Verify that the temperature is indeed 32°F using another thermometer that is known to be accurate. Then, simply place the sensor in the bath and adjust R6 until you get the correct meter reading.

HIGH-ACCURACY THERMOMETER



LINEAR TECHNOLOGY ***Fig. 89-5***

This circuit combines a current source and a platinum RTD bridge to form a complete high-accuracy thermometer. The ground-referred RTD sits in a bridge that is composed of the current drive and the LT1009 biased resistor string. The current drive allows the voltage across the RTD to vary directly with its temperature-induced resistance shift. The difference between this potential and that of the opposing bridge leg forms the bridge output. The RTD's constant drive forces the voltage across it to vary with its resistance, which has a nearly linear positive temperature coefficient. The nonlinearity could cause several degrees of error over the circuit's 0°C –400°C operating range.

The bridge's output is fed to instrumentation amplifier A3, which provides differential gain, while simultaneously supplying nonlinearity correction. The correction is implemented by feeding a portion of A3's output back to A1's input via the 10- to 250-kΩ divider. This causes the current supplied to R_P to slightly shift with its operating point, compensating sensor nonlinearity to within ±0.05°C. A1B, providing additional scaled gain, furnishes the circuit output. To calibrate this circuit, follow the procedure given in the diagram.

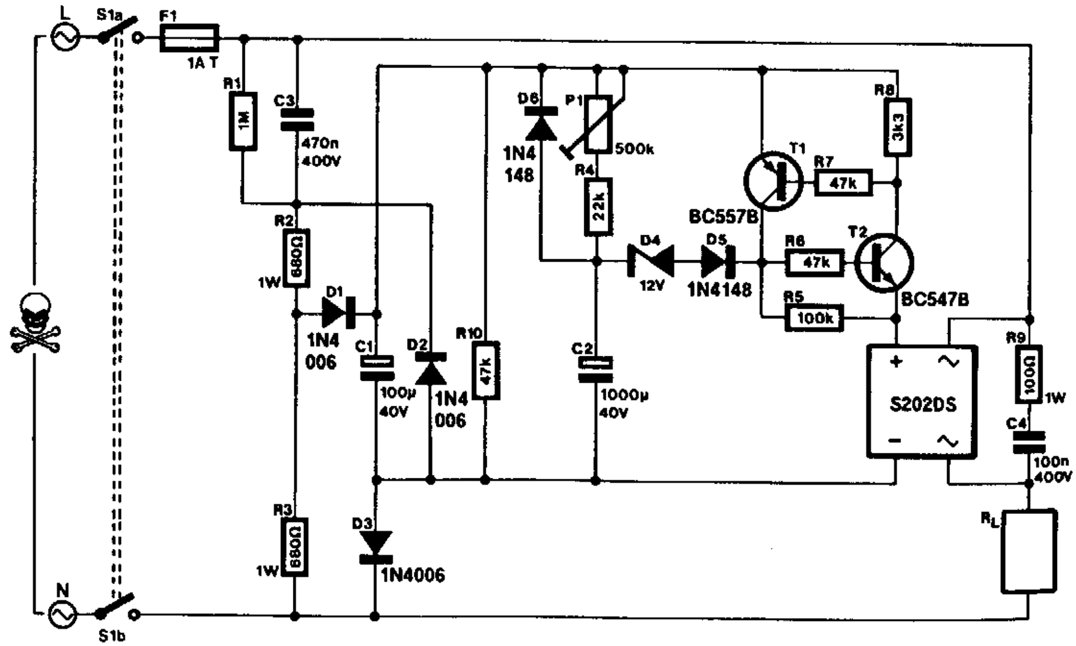
90

Timers

The sources of the following circuits are contained in the Sources section, which begins on page 679. The figure number in the box of each circuit correlates to the entry in the Sources section.

Mains-Powered Timer
Transmit Time Limiter
Long-Interval Programmable Timer
Programmable Timer for Long Intervals
Appliance Cutoff Timer
SCR Timer
Watchdog Timer/Alarm
10-Minute ID Timer
Adjustable Timer
Long-Duration Time Delay
Time-Out Circuit

MAINS-POWERED TIMER



ELEKTOR ELECTRONICS

Fig. 90-1

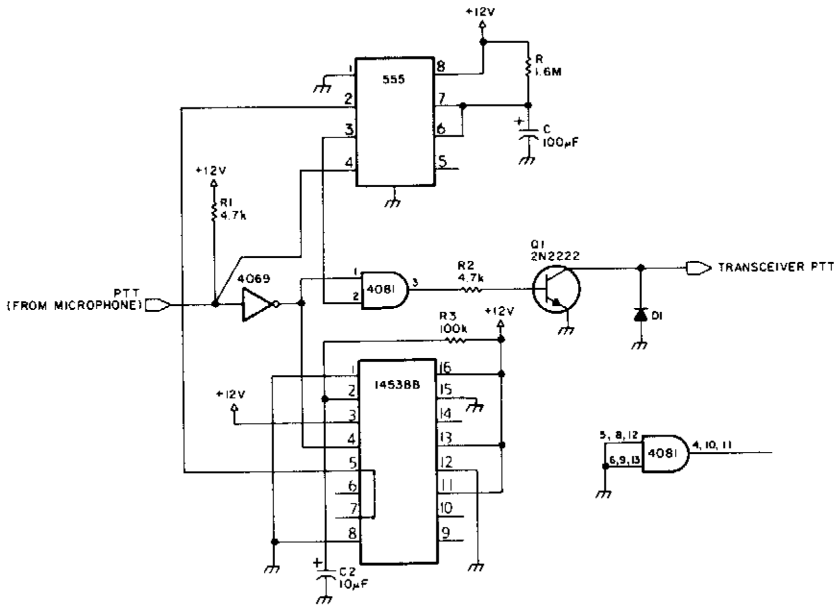
This timer can be inserted in a power line to provide a controllable delay before a load is energized. The mains voltage is reduced by C3 and rectified to give about 30 V across C1. This potential charges C2 slowly via R4/P1. When U_{C2} reaches about 14 V, electronic switch T1/T2 actuates a solid-state relay (a Sharp S202DS). When the mains voltage is removed, C2 discharges rapidly via D6 and R10. The delay extends from 15 s (P1 set to minimum resistance) to 5 min (P1 set to maximum resistance).

The solid-state relay needs cooling in accordance with the current drawn by the load: at up to 1 A, no heatsink is required; at 1 to 3 A (max), a 5×5 cm heatsink is advisable.

During the building of the circuit, consideration must be given to safety because many parts will be at mains potential. For instance, fitting the unit in an ABS or other man-made fiber enclosure is a must. If a potentiometer is used for P1, its spindle should be insulated. If a preset is used, it must not be accessible through a hole in the enclosure.

Switch S1 is a DPST that disconnects the circuit from the mains. Nevertheless, the only way to safely work on the circuit is to unplug the mains socket and allow C3 sufficient discharge time.

TRANSMIT TIME LIMITER



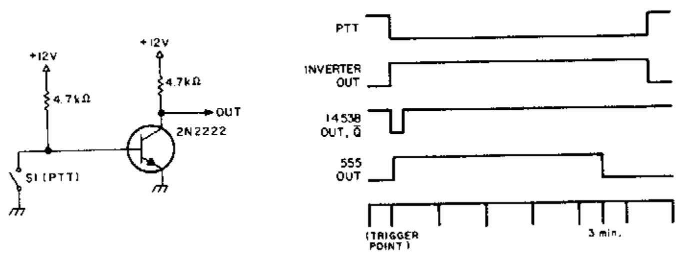
Values for Diverse Duty Cycles

T = 1.1 RC

T_s	R_Ω	C_f
180	1.6M	100µF
150	1.3M	100µF
120	1.1M	100µF
90	818.2k	100µF

180s = 3 min.
150s = 2 1/2 min.
120s = 2 min.
90s = 1 1/2 min.
(FOR 555 TIMER)

C>100µF TYPICALLY REQUIRES MORE EXPENSIVE LOW-LEAKAGE CAPACITORS.

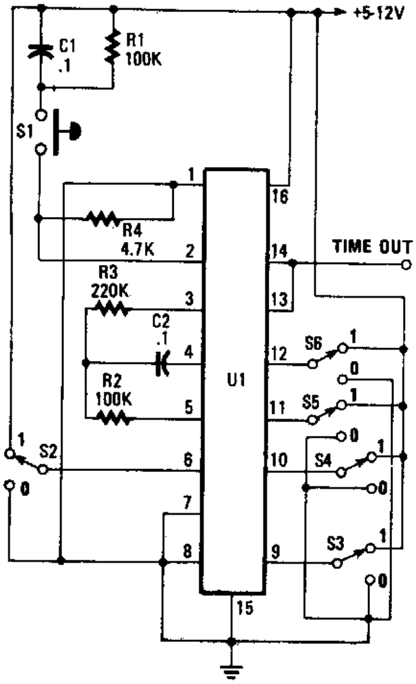


73 AMATEUR RADIO

Fig. 90-2

This circuit prevents making transmissions that are too long, which "time out" repeaters and/or tie up a communications channel too long. On transmit, the PTT (push to talk line) from the microphone is at ground. This causes one input of the AND gate (4081) to go high. The 555 is held in the reset mode. An MC1453B monostable multivibrator generates a pulse to the 555 and causes it to produce a gate of length approximately $1.1RC$, where R is selected for desired time delay and C is 100 µF. The output of pin 3 of the 555 causes the 4081 gate to go high, turns on Q1, and keys the transmitter. At the end of the cycle ($1.1RC$), the AND gate will lose one input, which turns off Q1 and unkeys the transmitter.

LONG-INTERVAL PROGRAMMABLE TIMER



BINARY CODE INPUT					
8 BYPASS PIN 6	D PIN 12	C PIN 11	B PIN 10	A PIN 9	SELECTED STAGE FOR OUTPUT
0	0	0	0	0	9
0	0	0	0	1	10
0	0	0	1	0	11
0	0	0	1	1	12
0	0	1	0	0	13
0	0	1	0	1	14
0	0	1	1	0	15
0	0	1	1	1	16
0	1	0	0	0	17
0	1	0	0	1	18
0	1	0	1	0	19
0	1	0	1	1	20
0	1	1	0	0	21
0	1	1	0	1	22
0	1	1	1	0	23
0	1	1	1	1	24

A

BINARY CODE INPUT					
8 BYPASS PIN 6	D PIN 12	C PIN 11	B PIN 10	A PIN 9	SELECTED STAGE FOR OUTPUT
1	0	0	0	0	1
1	0	0	0	1	2
1	0	0	1	0	3
1	0	0	1	1	4
1	0	1	0	0	5
1	0	1	0	1	6
1	0	1	1	0	7
1	0	1	1	1	8
1	1	0	0	0	9
1	1	0	0	1	10
1	1	0	1	0	11
1	1	0	1	1	12
1	1	1	0	0	13
1	1	1	0	1	14
1	1	1	1	0	15
1	1	1	1	1	16

B

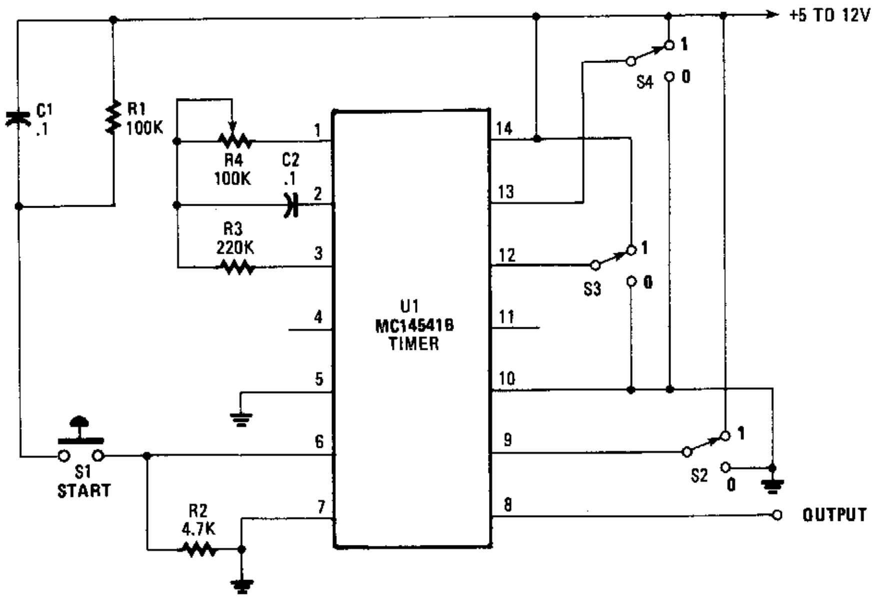
These truth tables can be used to set the desired output frequency. Refer to the text for details.

POPULAR ELECTRONICS

Fig. 90-3

Using an RC oscillator, an up to 24-stage ripple counter ($\div$) 16777216 or 2^{24}, and a 0.1-Hz count-rate with $R_2 = 39$ kΩ, $C_2 = 0.001$ μF, and $R_4 = 220$ kΩ for example, the count cycle would take about 654 s. This example shows the capabilities of this time circuit, using the Motorola MC14536 timer. A low-frequency oscillator can be used for longer time periods.

PROGRAMMABLE TIMER FOR LONG INTERVALS



COUNTER SELECTOR CHART

PIN 12	PIN 13	NUMBER OF COUNTER STAGES (N)	COUNT 2^N
0	0	13	8192
0	1	10	1024
1	0	8	256
1	1	16	65536

The inputs at pins 12 and 13 of the MC14415 programmable counter determine the number of counter stages selected, and therefore the count.

POPULAR ELECTRONICS

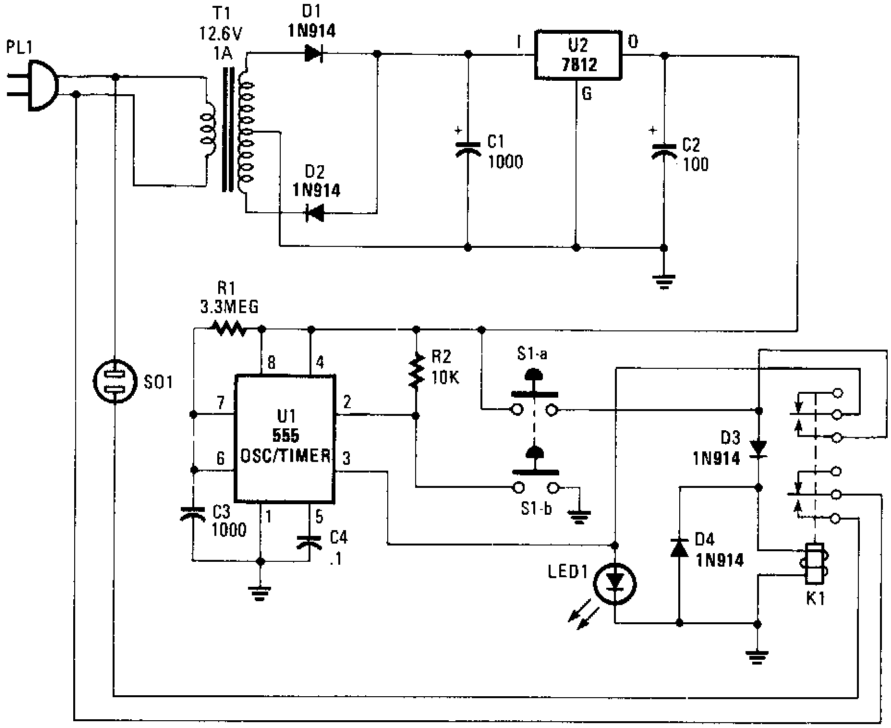
Fig. 90-4

By using an RC oscillator and a programmable divider, this counter can run for hours. An interval oscillator runs at a frequency given by (see figure schematic):

$$f = \frac{1}{2.3\,R_4C_2} \text{ and } R_3 \approx 2R_2$$

By using, for example, $R_4 = 390$ kΩ and $C_2 = 10$ μF and R_2, the oscillator can run at 0.1 Hz. Divided by 65536, this is a cycle of approximately 655 000 s (182 hours, slightly more than a week).

APPLIANCE CUTOFF TIMER

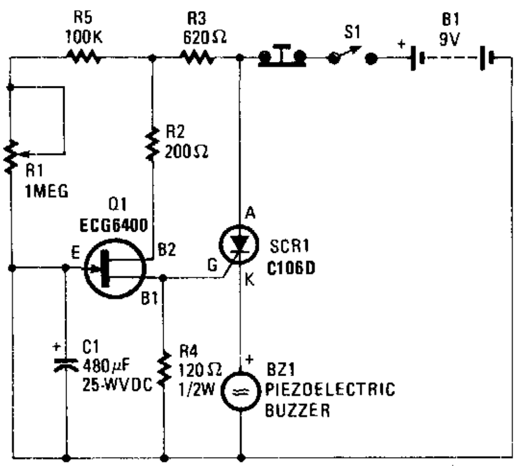


POPULAR ELECTRONICS

Fig. 90-5

Suitable for cutting off an appliance or other ac load, this timer will cut the ac power after a period determined by R1/C3, as shown, for about 40 minutes. K1 is a relay that should handle about 10 A. S1A and S1B is a momentary switch that starts the timer cycle.

SCR TIMER

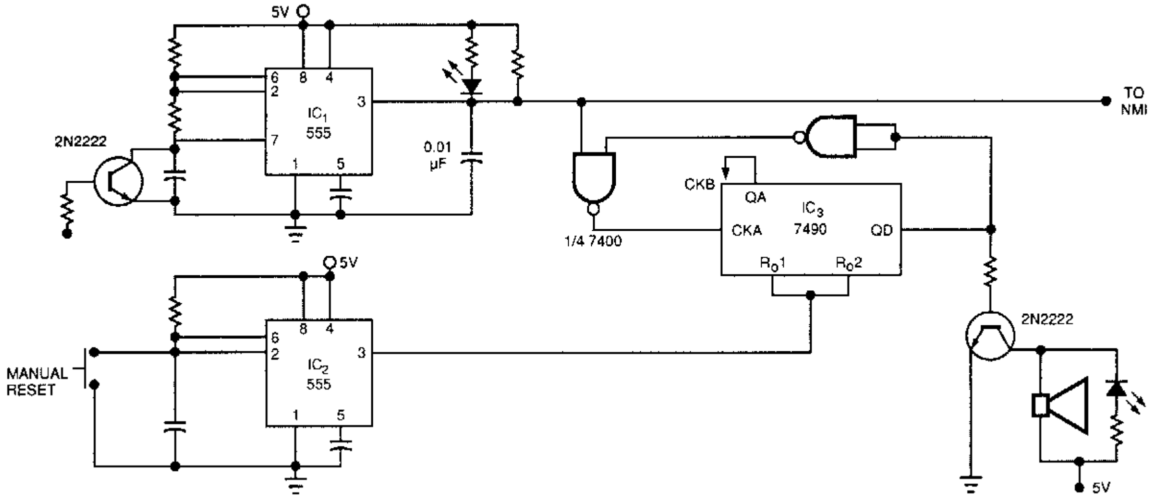


Depending on the R1 (adjustable) C1 time constant, when C1 charges up to a certain level, Q1 conducts, triggers SCR1, and sounds BZ1. A pushbutton resets the circuit.

HANDS-ON ELECTRONICS

Fig. 90-6

WATCHDOG TIMER/ALARM

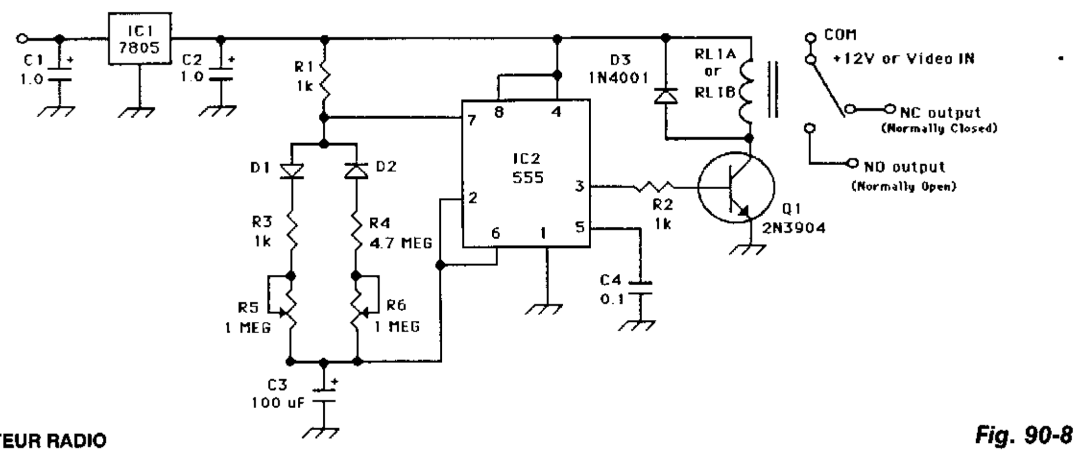


EDN ***Fig. 90-7***

The watchdog timer contains a counter, IC3, in addition to the usual retriggerable 555 timer, IC1. The counter will sound an audible alarm if the watchdog timer trys to reset the µP a certain number of times (8, in the case of the counter). The alarm indicates that despite numerous resets, the system µP has failed to restart successfully, and the system is truly dead.

A second 555 timer, IC2, resets the counter, 1C3, for the duration of the manual system restart. The design could be modified so that system µP resets the counter.

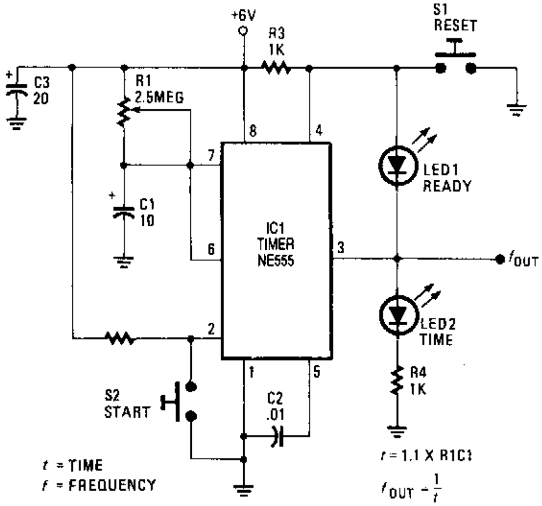
10-MINUTE ID TIMER



73 AMATEUR RADIO ***Fig. 90-8***

Designed to automatically identify a transmitter every 10 minutes, this 555 circuit has adjustable charge and discharge paths. The IC should be a standard 555 type, not a CMOS type. C3 should be tantalum. The relay is a small 5-V reed type.

ADJUSTABLE TIMER

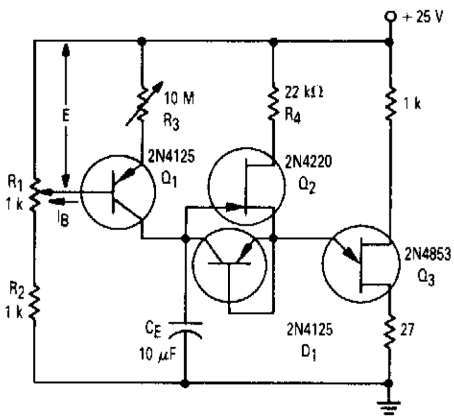


HANDS-ON ELECTRONICS

Fig. 90-9

LEDs indicate at a glance what the status of the circuit is at any given moment. Once the reset switch, S1, makes contact, the timer remains in that state until the start switch, S2, is pressed. When either switch is activated, LED1 (ready) and the time indicator, LED2, keep track of the situation. Although not necessary, the two LEDs should be of different colors (for example, red for "ready" and green for "time").

LONG-DURATION TIME DELAY



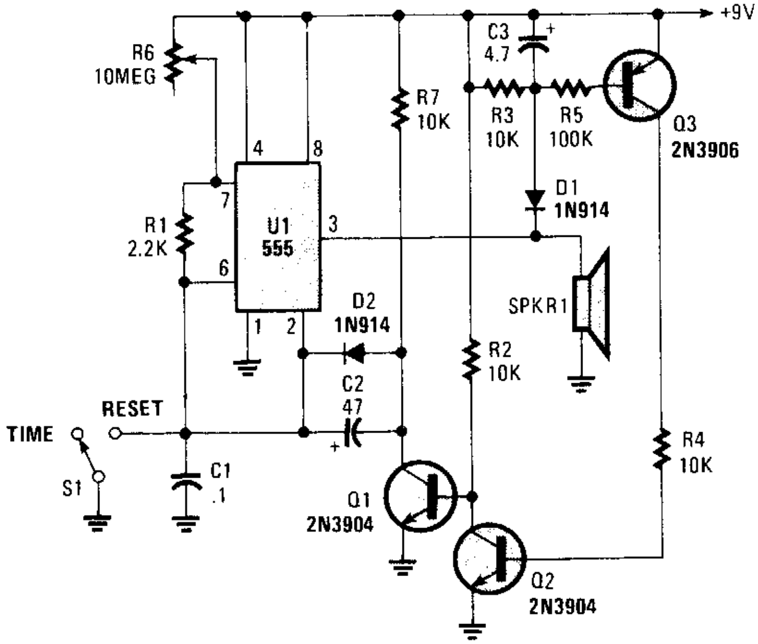
MOTOROLA

Fig. 90-10

Transistor Q1 and resistors R1, R2, and R3 form a constant current source and the charge current might be adjusted to be as low as a few nanoamperes. This current would, of course, not be sufficient to fire the UJT where $I_P = 0.2\ \mu A$, unless the peak current was supplied from another source. Field-effect transistor Q2, acting as a source follower, supplies the current flowing into the emitter lead prior to firing and diode D1 provides a low-impedance discharge path for C_E. D1 must be selected to have a leakage that is much lower than the charge current.

Because I_B is small, the delay time will vary linearly with R3. The voltage (E), applied across R3 and the base-emitter junction of Q1, is set by the variable resistor R1. Time delays up to 10 hours are possible with this circuit. Resistor R4, in series with the FET drain terminal, must be large enough not to allow currents in excess of I_V to flow when the UJT is on. Otherwise, the UJT will not turn off and the circuit will latch up.

TIME-OUT CIRCUIT



POPULAR ELECTRONICS

Fig. 90-11

This circuit operates in the astable mode and at the end of the first period (up to several minutes), it produces a tone. When S1 is placed in the time position, Q3 is cut off because pin 3 of U1 is high and D1 holds Q3 in cutoff. Q2 is off, and Q1 is on, which grounds the negative end of D2 and C2. Therefore, C1 and C2 are returned to ground.

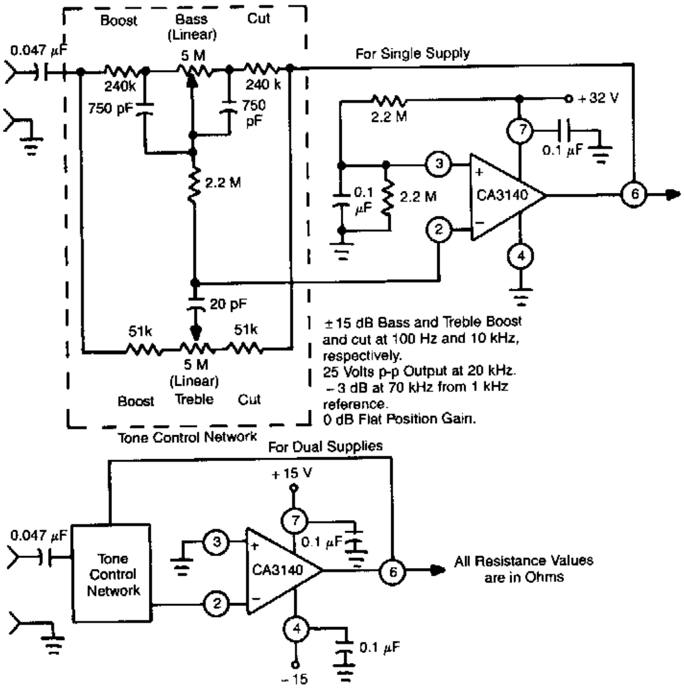
After a time of about 1.1 R_6 $(C_1 + C_2)$, the timer cycle completes and pin 3 U1 goes low. This turns on Q3 and Q2, cuts off Q1, and effectively disconnects C2. Now, the circuit oscillates with a period determined by R7 and C1, because D2 is forward-biased. A tone is then generated and can be heard from SPKR1. Closing S2 resets the circuit.

91

Tone-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 680. The figure number in the box of each circuit correlates to the entry in the Sources section.

BAXANDALL TONE-CONTROL AUDIO AMPLIFIER

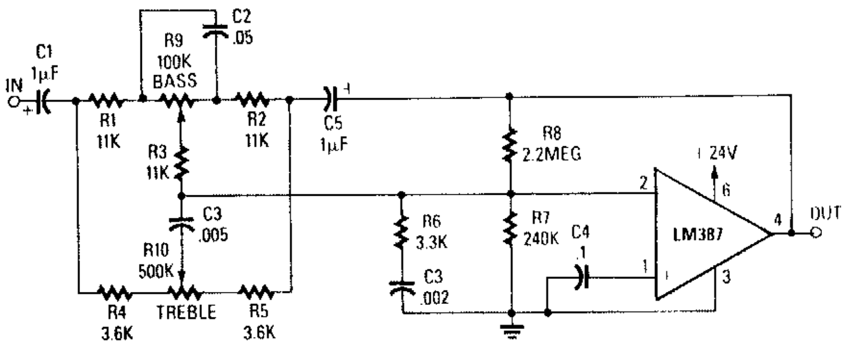


GE/RCA

Fig. 91-1

This circuit exploits the high slew rate, high input impedance, and high output-voltage capability of CA3140 BiMOS op amp. It also provides mid-band unity gain with standard linear potentiometers.

ACTIVE TONE CONTROL

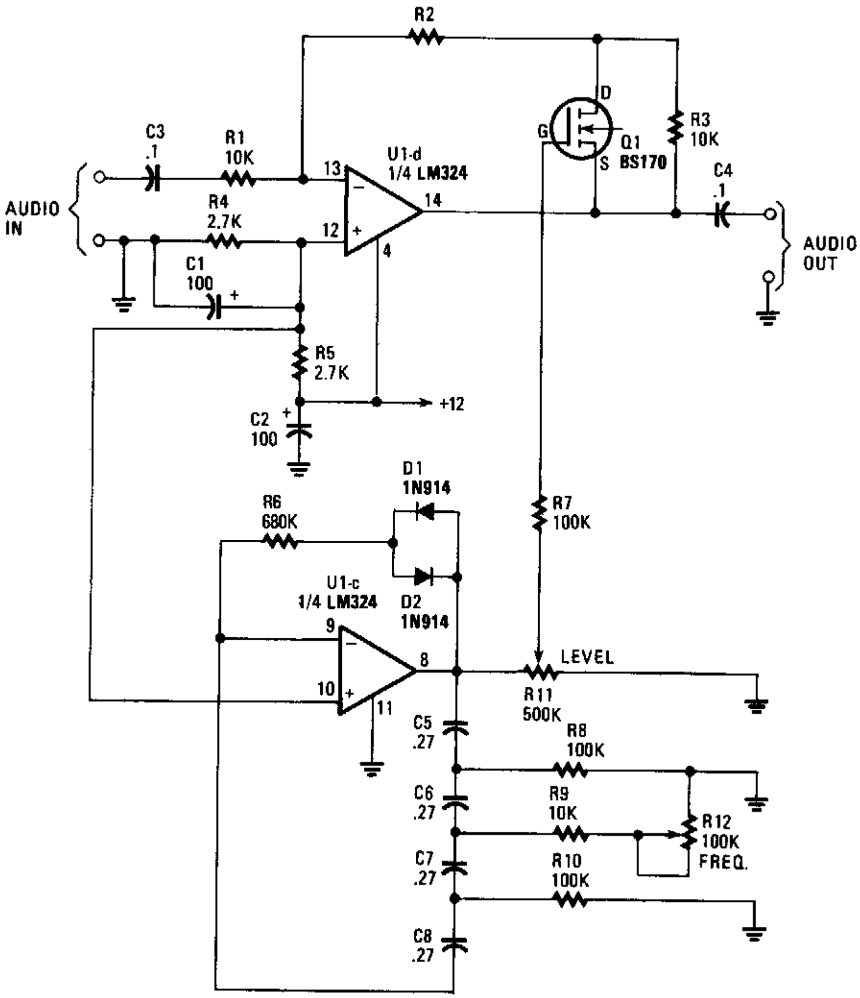


RADIO-ELECTRONICS

Fig. 91-2

The use of a low noise LM387 in a feedback circuit provides 20-dB boost or rejection of treble and bass. The supply voltage is +24 V.

TREMOLO CIRCUIT



HANDS-ON ELECTRONICS

Fig. 91-3

This circuit adds a VLF AM component to an audio signal. This effect is widely used in musical instruments. U1C, a phase-shift oscillator operating at a few Hz applies a signal to Q1, which modulates the gain of U1D. R11 varies the level of the effect, while R12 varies the frequency.

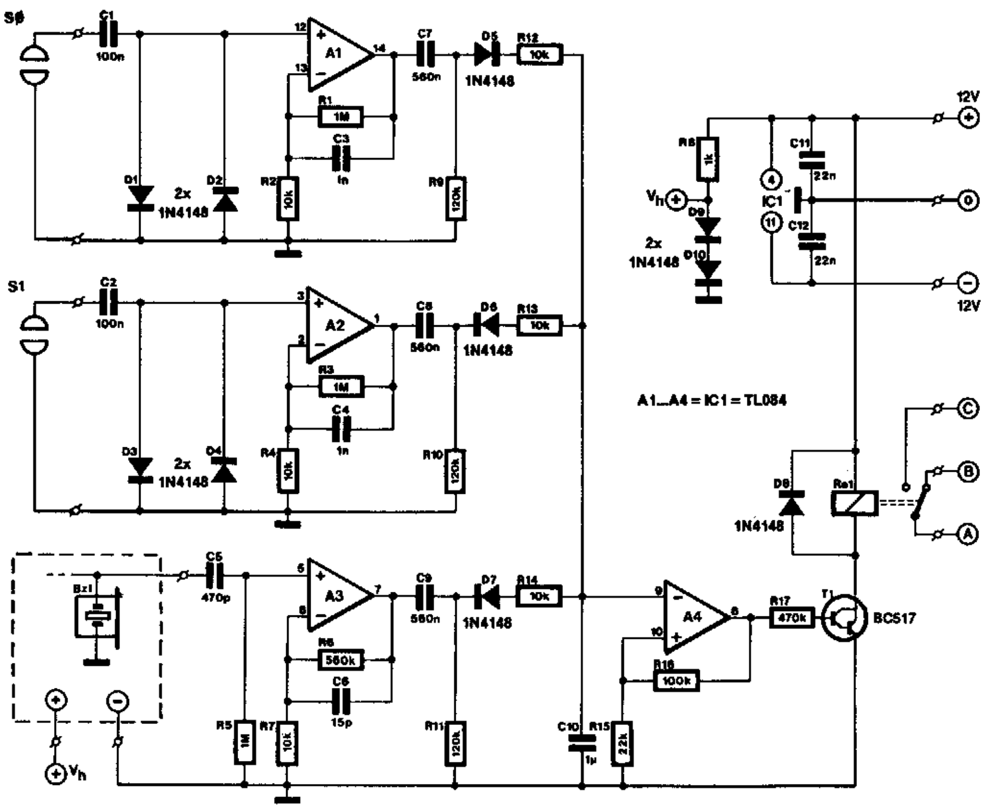
92

Touch Controls

The sources of the following circuits are contained in the Sources section, which begins on page 680. The figure number in the box of each circuit correlates to the entry in the Sources section.

Sensor Switch and Clock
Touch Switch I
On/Off Touch Switch
Touch Switch II
Hum-Detecting Touch Sensor
Time-On Touch Switch

SENSOR SWITCH AND CLOCK



ELEKTOR ELECTRONICS *Fig. 92-1*

One TL084 IC and an old quartz watch enable the construction of a deluxe on/off switch. Two of the four op amps contained in the TL084 (A1 and A2) are used to amplify the input signals from the sensors by one hundredfold (with the component values as shown in the diagram). Just touching the sensors with a finger causes a good 50-Hz input signal (hum). Notice that the amplification drops rapidly with rising frequency.

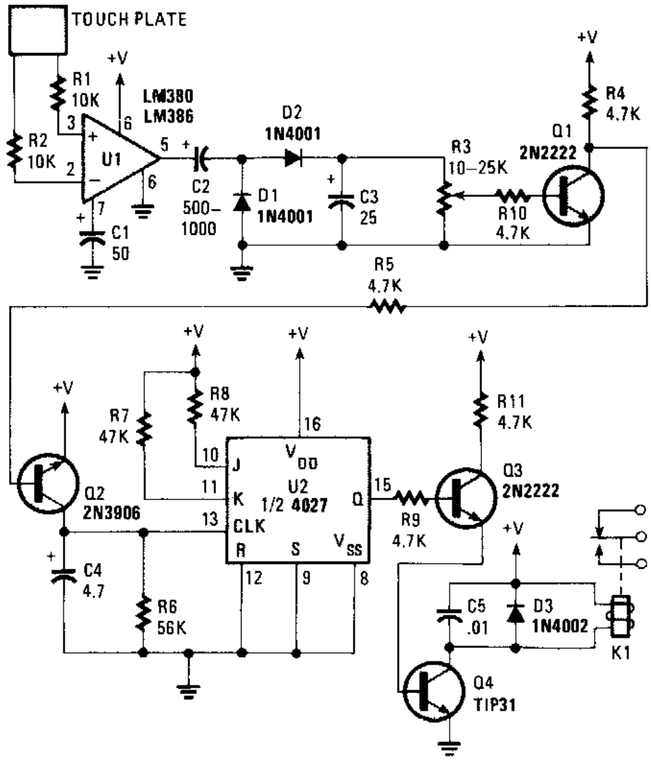
Diodes D5 and D6 rectify (single-phase) the 50-Hz signal. Because the diodes are connected in antiphase, touching the "off" sensor causes a positive potential across C10, whereas touching the "on" sensor produces a negative potential across C10.

Op amp A4 is connected as an inverting bistable so that a negative potential across C10 causes relay Re1 to be energized. Because of feedback resistor R16, this state is maintained until the other sensor is touched.

The relay can also be energized at a predetermined time with the aid of a quartz watch. The 1.2-V supply for the watch is derived from the voltage drop across diodes D9 and D10; it can be increased to 1.8 V by adding a third diode.

The piezo buzzer in the watch is connected to the input of A3 via C5. As soon as the alarm goes off (the hour signal must be off), the voltage across C10 becomes negative, the relay is energized, and the load is switched on. The circuit, excluding the relay, draws a current of about 20 mA.

TOUCH SWITCH I



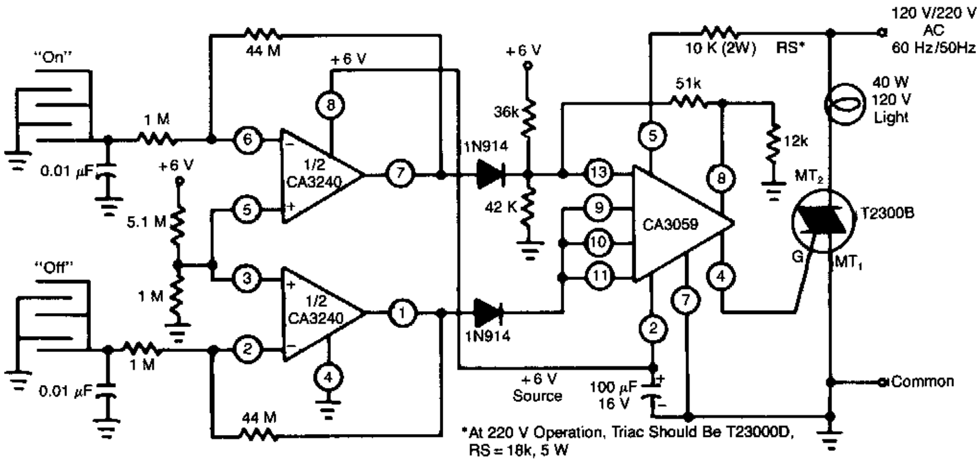
POPULAR ELECTRONICS *Fig. 92-2*

This switch reacts to the touch of a finger to turn lights and/or appliances on or off. The device uses the human body as an antenna to pick up 60-Hz hum, which is applied to a metal plate by your finger. The signal is fed to the input of U1, an LM380 audio-power amplifier. An LM386 should work as well.

The 60-Hz output from the amplifier is rectified by D1 and D2, then filtered by C3. Potentiometer R3 sets the trigger voltage used to saturate Q1. When Q1 turns on, the collector end of R4 goes almost to ground and provides the needed voltage to turn on Q2. Transistor Q2 turns on and clocks. The flip-flop is configured for toggle-mode operation, so its output switches states with each clock pulse.

The 4027 (U2) is wired to toggle by tying the J and K inputs high and the set and resets low. Transistor Q3 is connected to the *Q* output through the 4.7-kΩ resistor. Transistor Q3 drives Q4, the relay driver. Be sure that the load does not exceed the relay ratings.

ON/OFF TOUCH SWITCH

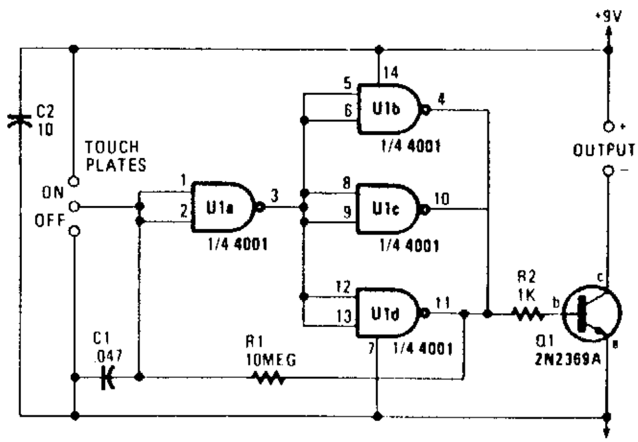


GE/RCA

Fig. 92-3

This circuit uses a CA3240 dual BiMOS op amp to sense small currents flowing between the contact points on a touch plate. The high input impedance of the CA3240 allows the use of 1-MΩ resistors in series with the touch plates to ensure user safety. A positive output on either pin 7 (ON) or pin 1 (OFF) of the CA3240 actuates the CA3059 zero-voltage switch, which then latches the triac on or turns it off. The internal power supply of the CA3059 powers the CA3240.

TOUCH SWITCH II

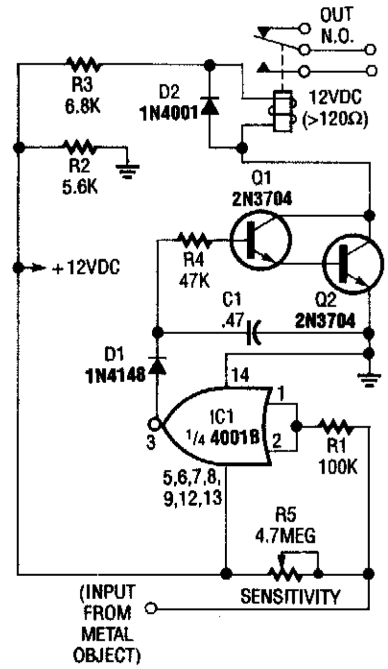


WELS' THINK TANK

Fig. 92-4

U1A and U1B/U1C/U1D form a bistable multivibrator that drives Q1, which switches the load. Touching the two upper contacts makes Q1 conduct; the two lower contacts cause Q1 to cut off.

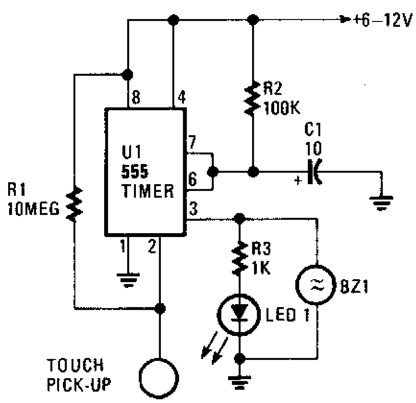
HUM-DETECTING TOUCH SENSOR



RADIO-ELECTRONICS *Fig. 92-5*

This touch sensor uses the 60-Hz hum pickup by the human body to drive a detector and relay drivers, Q1 and Q2. R5 controls sensitivity of the circuit.

TIME-ON TOUCH SWITCH



HANDS-ON ELECTRONICS *Fig. 92-6*

The circuit is built around a 555 oscillator (U1), which is turned on when a trigger is applied by touching the touch terminal to pin 2 of U1. When activated, LED1 and BZ1 (a piezoelectric buzzer) turn on for the time period set by the values of R_2 and C_1. The ON-time of the touch circuit can be altered by changing the values of C_1 and R_2.

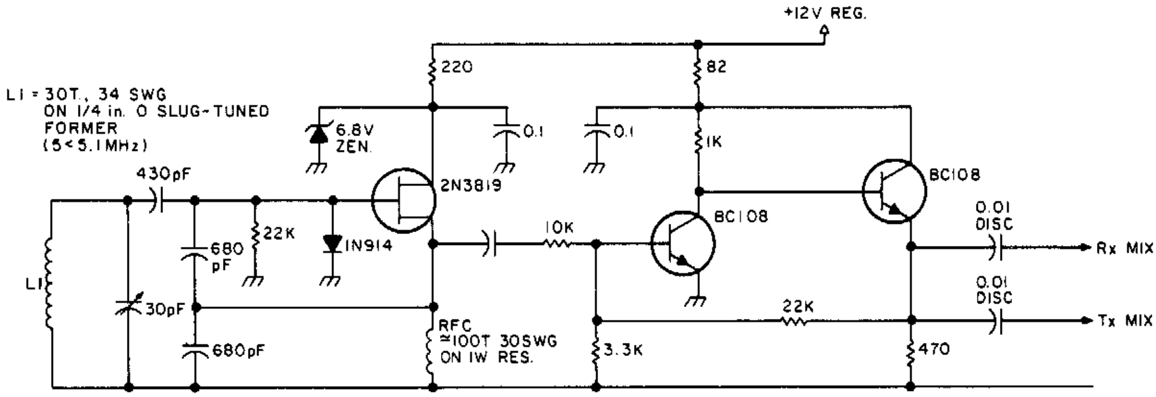
This touch switch can be powered from batteries so that it need not be near a 60-Hz power source for triggering. The extremely small amount of current supplied to the trigger input through the 10-MΩ resistor. R1, makes the input circuitry very sensitive to any external loading, and it is easily triggered by touching the pickup.

93

Transmitters and Transceivers

The sources of the following circuits are contained in the Sources section, which begins on page 680. The figure number in the box of each circuit correlates to the entry in the Sources section.

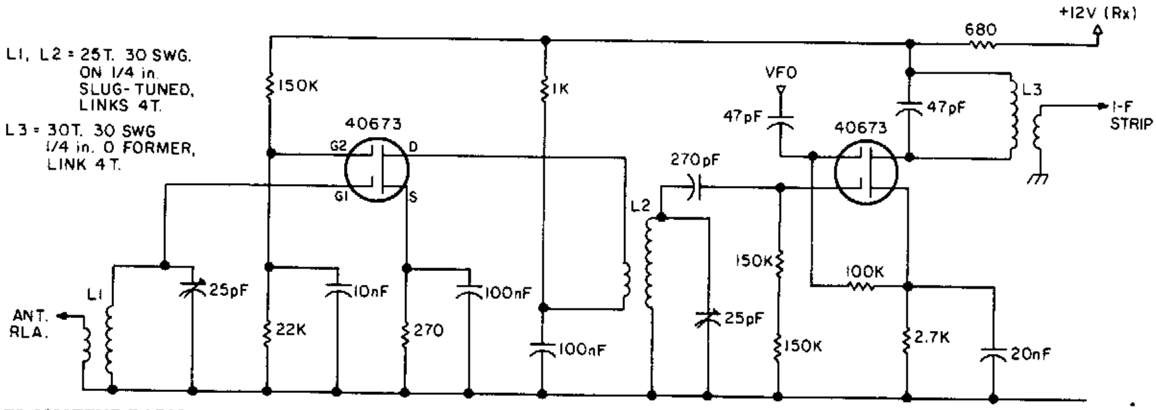
20-m CW TRANSCEIVER



73 AMATEUR RADIO

Fig. 93-1(a)

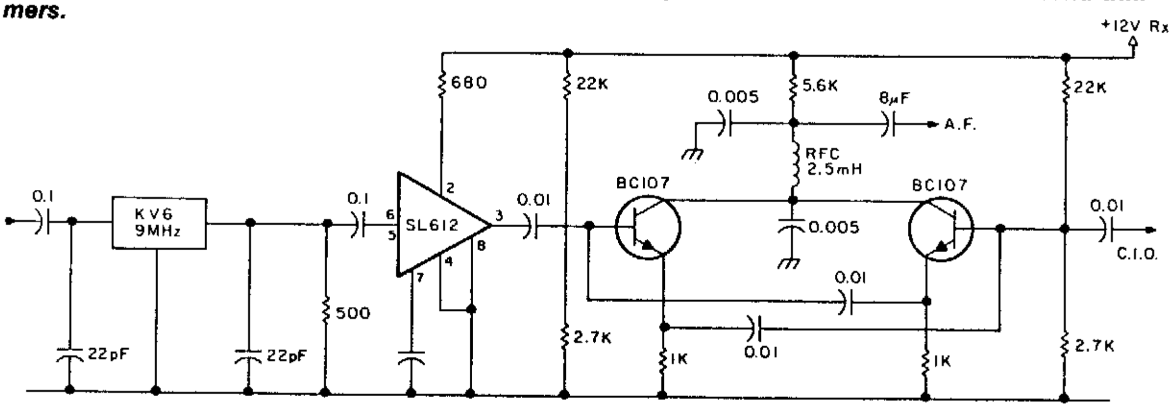
Colpitts VFO circuit.



73 AMATEUR RADIO

Fig. 93-1(b)

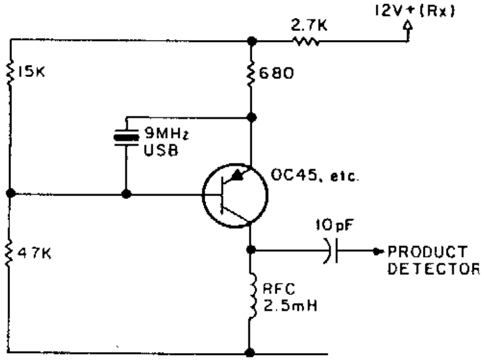
The receive front-end. The associated tuned circuits are peaked to the center of the CW band trimmers.



73 AMATEUR RADIO

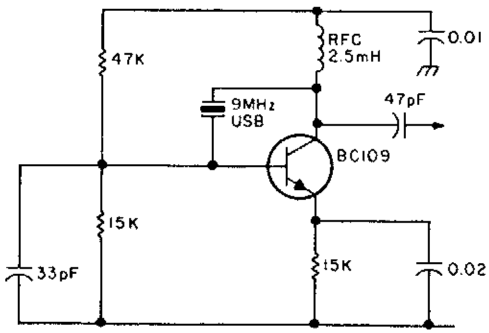
Fig. 93-1(c)

Receive filter, IF, and product detector.



73 AMATEUR RADIO

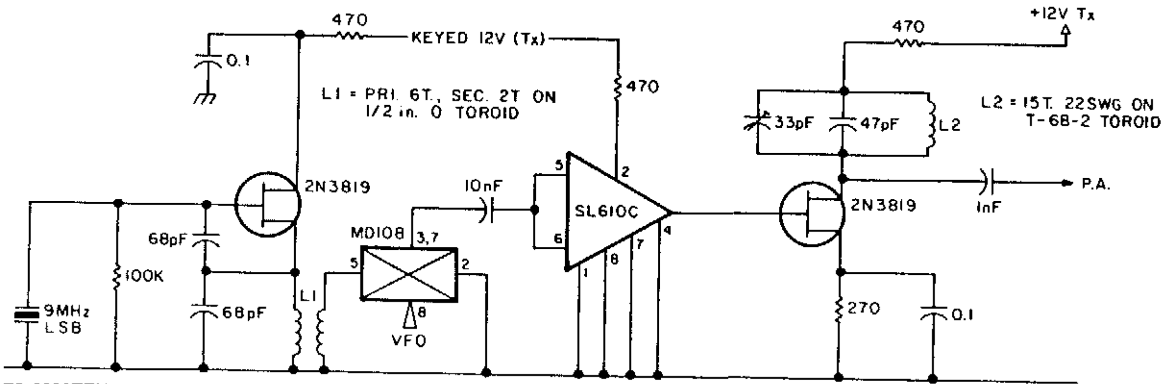
Fig. 93-1(d)



73 AMATEUR RADIO

Fig. 93-1(e)

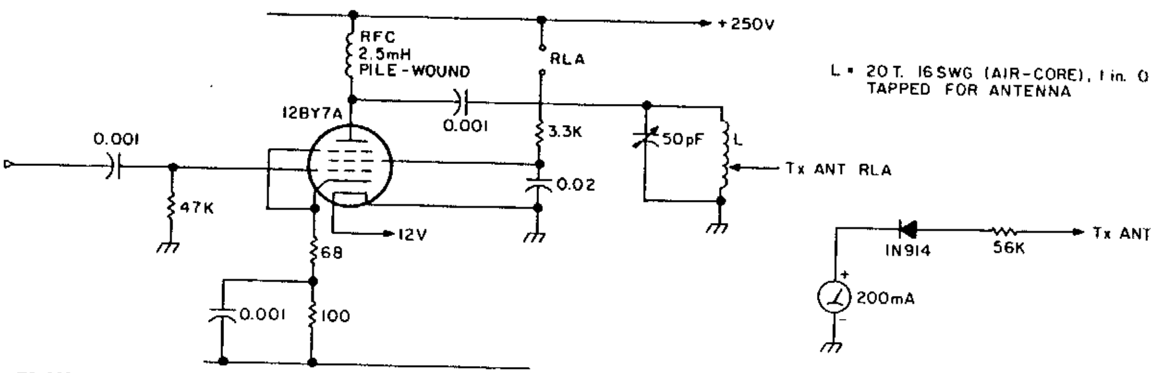
What's in the junk-box? Pick between two receive carrier circuits.



73 AMATEUR RADIO

Fig. 93-1(f)

Transmit mixer and predriver.

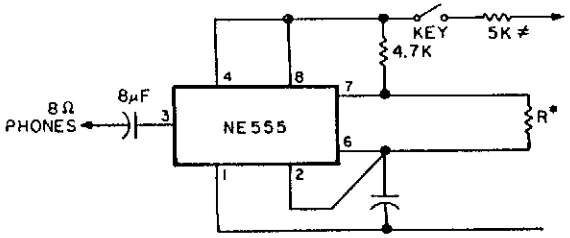


73 AMATEUR RADIO

Fig. 93-1(g)

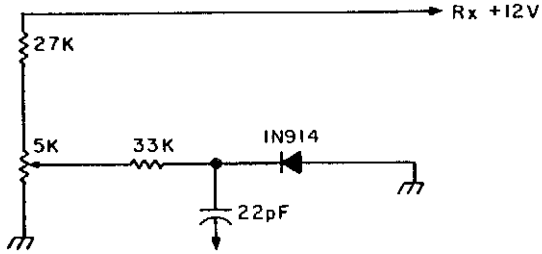
A simple diode/meter circuit to measure relative power output.

20-m CW TRANSCEIVER (*Cont.*)



R* : SELECT FOR DESIRED TONE (10<50K)
R≠ : VARY INPUT RES. FOR REQUIRED OUTPUT.

73 AMATEUR RADIO *Fig. 93-1(h)*



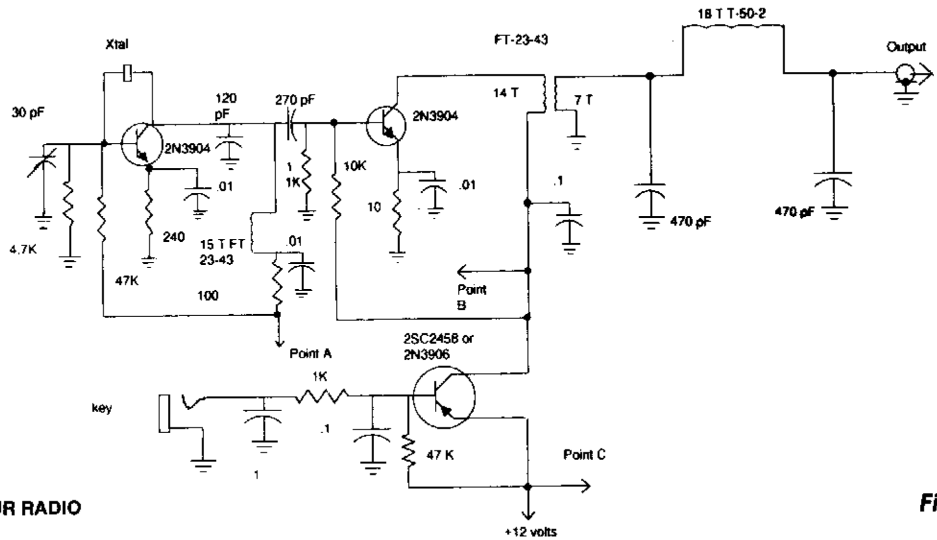
73 AMATEUR RADIO *Fig. 93-1(i)*

The Colpitts VFO circuit (Fig. 93-1(a)) has gate clamping to improve stability. It is followed by two buffers: the second provides individual outputs to the transmitter and receiver. An RIT circuit operates on receive. The front end uses 40673 dual-gate MOSFETs (Fig. 93-1(b)). The tuned circuits are peaked to the center of the CW band with the trimmer. The mixer output is link-coupled to a KVG 9-MHz SSB filter whose output is amplified by an SL612 IF amplifier IC (Fig. 93-1(c)).

The product detector uses two BC 107 transistors. Carrier reinjection is from a crystal oscillator using the USB crystal supplied with the KVG filter. Figure 93-1(d) and 93-1(e) contain: an FET oscillator and a pnp bipolar oscillator. Use either or alter the circuit polarities to suit an npn transistor. The transmitter mixer is an MD108 (Fig. 93-1(f)), fed from the VFO and the LSB carrier-injection oscillator.

A simple diode/meter circuit measures the relative power output (Fig. 93-1(g)). Sidetone is provided by an NE555 circuit (Fig. 93-1(h)).

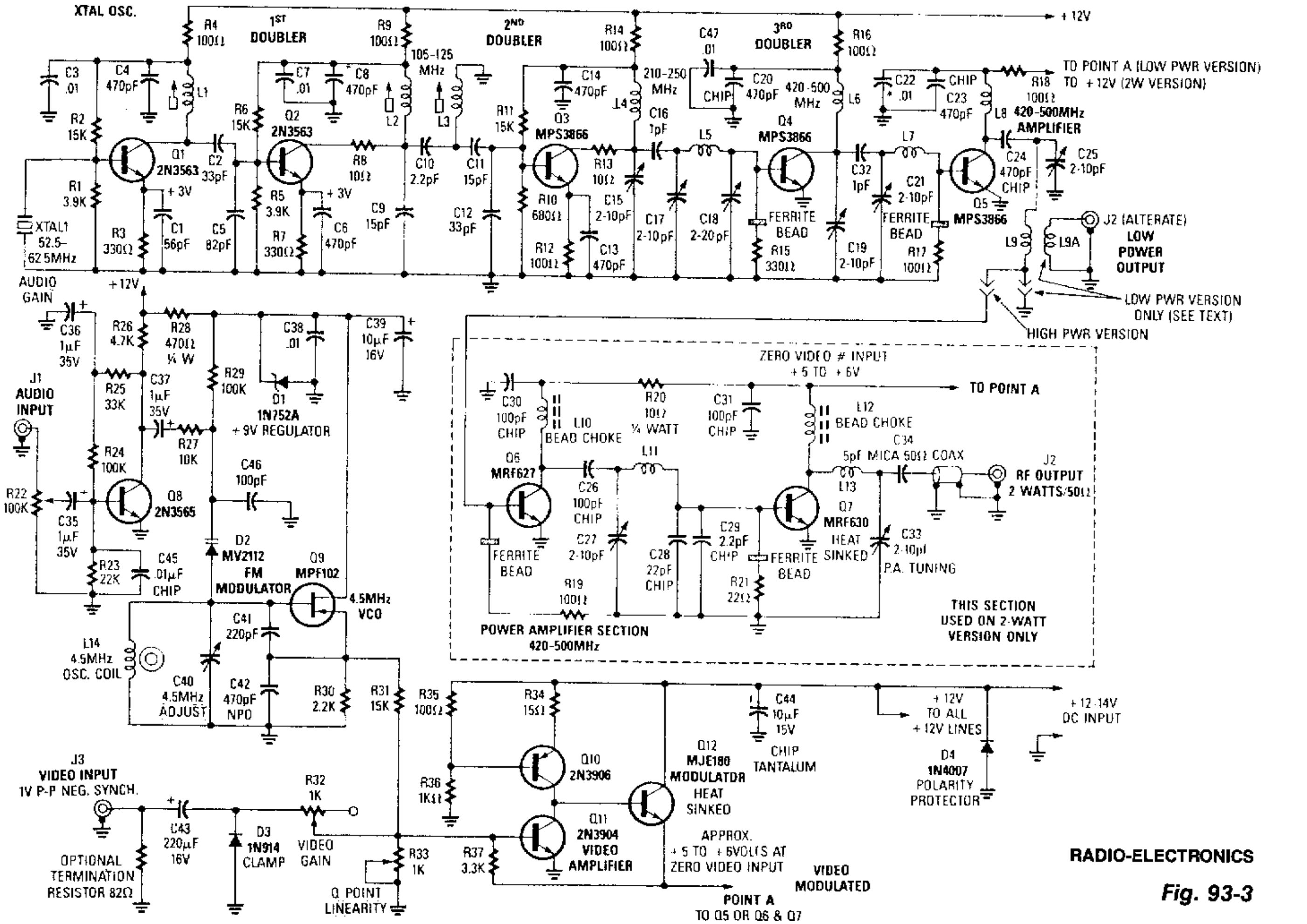
LOW-POWER HF TRANSMITTER



73 AMATEUR RADIO *Fig. 93-2*

This transmitter runs up to 1/2 W on the 40-m amateur band. Coils are wound on T25 and FT23 toroids, respectively. Point A is connected to point B to key the oscillator, along with the final (point C) to leave the oscillator running constantly. Use 1/10-W resistors and VERY small components.

AMATEUR TELEVISION TRANSMITTER

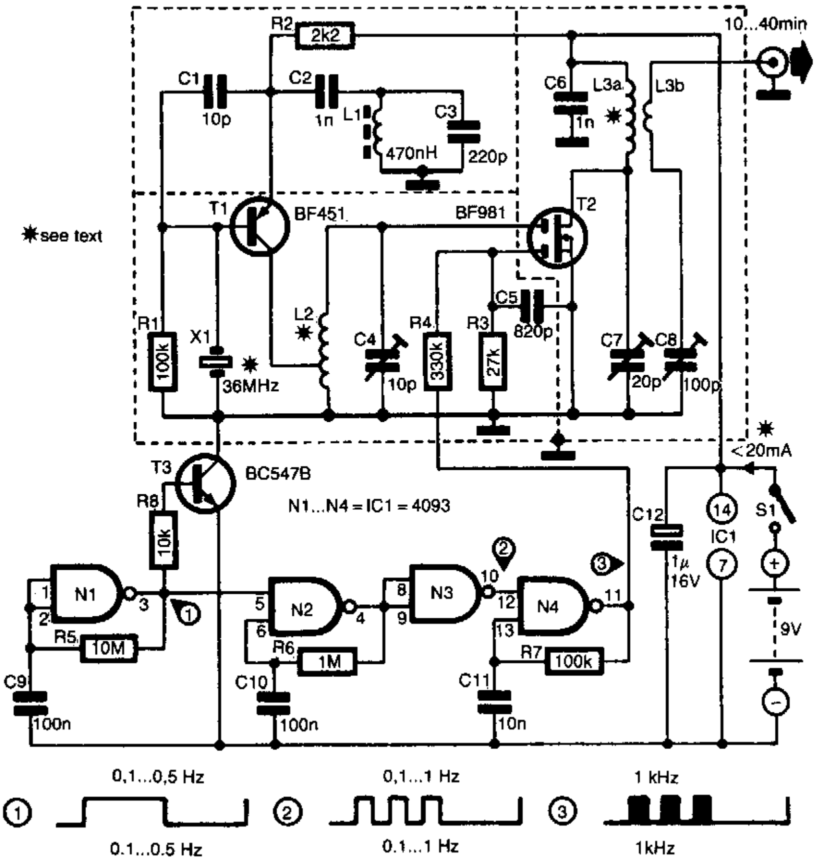


RADIO-ELECTRONICS

Fig. 93-3

Although the unit was designed for 2 W operation, Q6 and Q7 stages and associated components can be omitted and about 1 to 30 mW of RF can be obtained by link coupling to L9. A complete set of parts, including PC board is available from North Country Radio, P.O.Box 53, Wykagyl Station, New Rochelle, NY 10804.

2-m TRANSMITTER



ELEKTOR ELECTRONICS

Fig. 93-4

The transmitter was designed specifically for use by radio amateurs as a radio beacon. As such, it provides a good-quality signal that is free of unwanted harmonics.

Transistor T1, in association with crystal X1, operates as a 36-MHz oscillator. Filter L1/C3 obviates any tendency of the circuit to oscillate at 12 MHz (the fundamental frequency of the crystal).

Circuit L2/C4 is tuned to the fourth harmonic of the oscillator signal (144 MHz). This signal is fed to the aerial via a buffer stage that consists of T2, a double-gated FET. The (amplitude) modulating signal is applied to the second gate of the buffer. The output power of the transmitter has been kept low, about 10 to 40 mW.

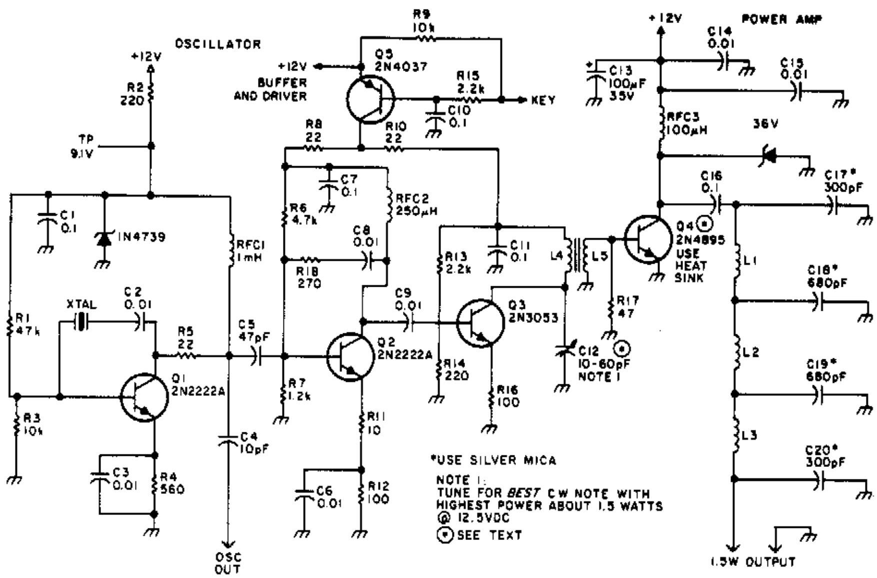
The modulating signal is generated by N1, an oscillator that switches the transmitter on and off via transistor T3. The switching rate lies between 0.1 and 0.5 Hz. When the output of N1 is low, T3 is switched off, and the transmitter is inoperative because the supply is disabled. When the output of N1 is high, T3 is on and the transmitter operates normally.

The digital pattern at the gate of T2 shapes the modulating signal. Gate N2 generates a square wave at a frequency of 0.1 to 1 Hz. As long as the output of T3 is high, N4 oscillates at a frequency of about 1 kHz. At the relevant gate of N2, there is, therefore, a periodic burst-signal at 1 kHz, and this signal is used to modulate the transmitter.

The digital pattern at the relevant gate of T2 can be varied to individual requirements by altering the values of the feedback resistors in the digital chain. The transmitter is calibrated by setting trimmers C4, C7, and C8 for maximum output power.

Inductors L2 and L3 are wound from 0.8-mm diameter enamelled copper wire: L2 = 5 turns with a tap of 1 turn from ground; L3A = 3 turns and L3B = 2 turns. The coupling between L3A and L3B should be arranged for maximum output power. The circuit draws a current of only 20 mA, which enables the transmitter to be operated from a 9-V battery for several hours.

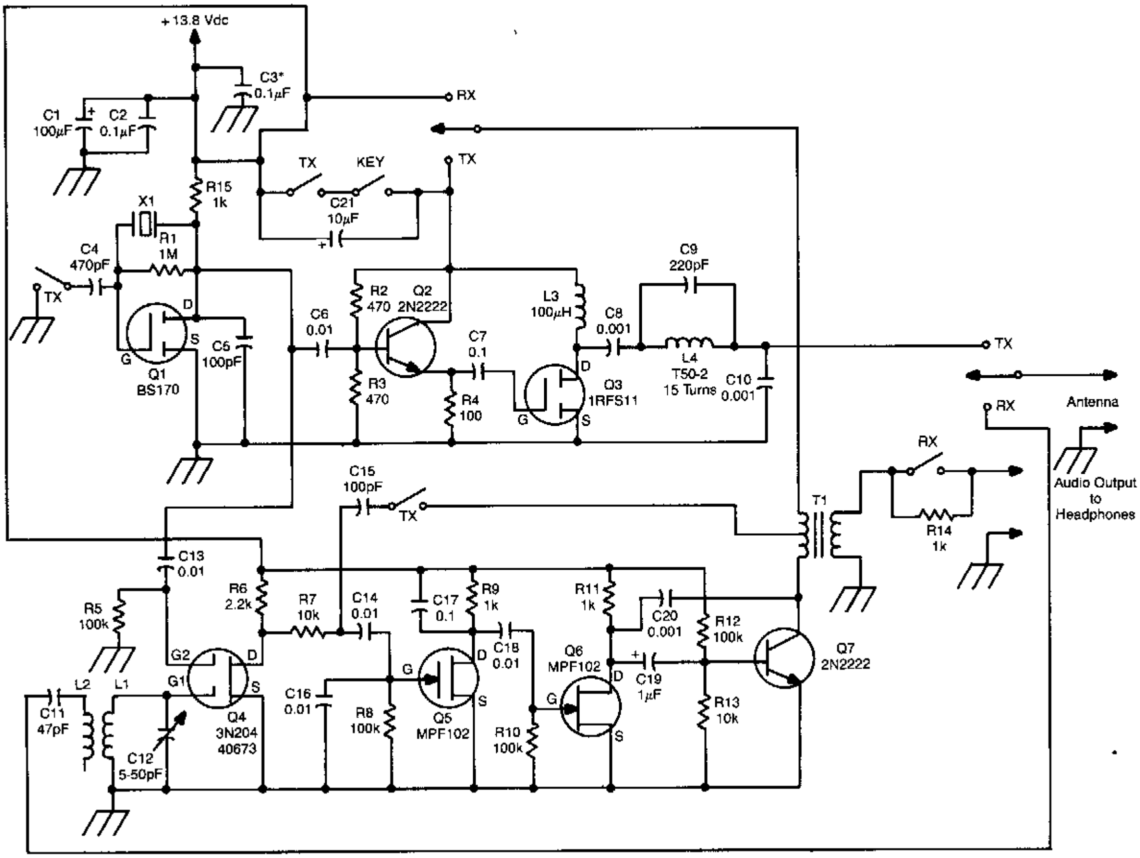
HF LOW-POWER CW TRANSMITTER



73 AMATEUR RADIO *Fig. 93-5*

Suitable for amateur use, this 1.5-W transmitter runs on a 12-V supply. Q1 is an oscillator using a surplus FT243 crystal. Q2 is a buffer driver and is keyed via keying transistor Q5. Q3 acts as a driver for Q4 (which should be heatsinked). Q4 develops about 1.5-W output. Coil data is given in the parts list. C12 is adjusted for best power output.

5-W 80-m CW TRANSCEIVER

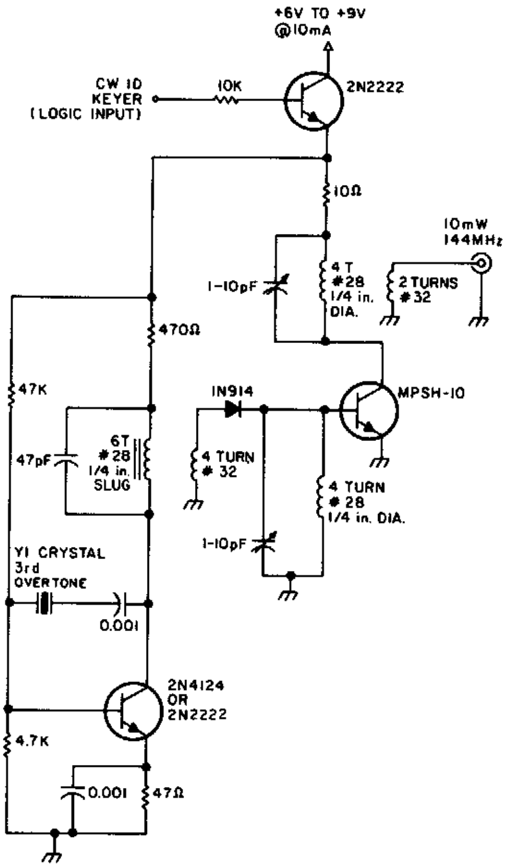


73 AMATEUR RADIO

Fig. 93-6

This transceiver has a 3-stage transmitter and a direct-conversion receiver. Q1 is the transmitter's oscillator, and the frequency is controlled by X1, which also serves as the receiver local oscillator. Buffer Q2 drives final amplifier Q3 to about 5 W output. The B+ lead to these stages is keyed. The receiver consists of mixer Q4 followed by high gain amplifiers Q5/Q6/Q7. The audio signal appears at the secondary of Q7. In the transmit mode, Q5/Q6/Q7 serve as a sidetone oscillator. A 6PDT switch is required for the T/R switching.

LOW-COST BEACON TRANSMITTER



73 AMATEUR RADIO ***Fig. 93-7***

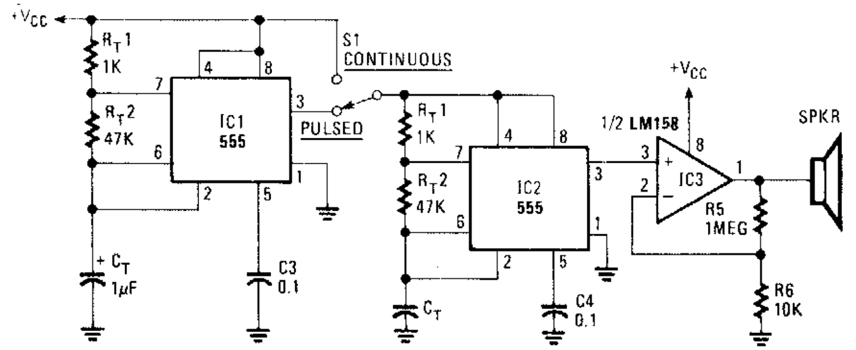
This transmitter uses a 48-MHz crystal oscillator to drive a diode tripler to the 144-MHz amplifier. The output is 5 to 10 mW.

94

Ultrasonic Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 680. The figure number in the box of each circuit correlates to the entry in the Sources section.

ULTRASONIC SOUND SOURCE



RADIO-ELECTRONICS

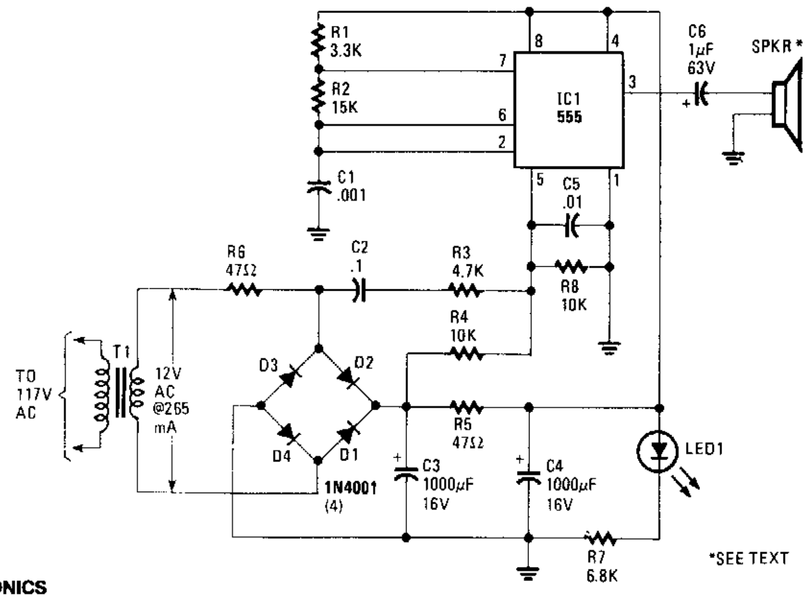
Fig. 94-1

Using two NE555 timer IC devices, this circuit generates either pulsed or continuous ultrasonic signals. The sound frequency is:

$$f = \frac{1.44}{C_T(R_{T1} + R_{T2})}$$

The values of C_T for both pulse rate and ultrasonic frequencies can be calculated this way. SPKR is a small hi-fi tweeter.

ULTRASONIC PEST REPELLER I

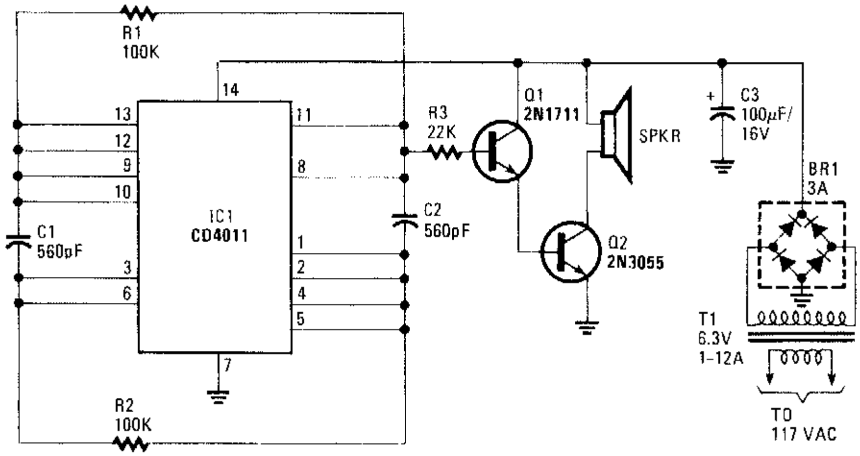


RADIO-ELECTRONICS

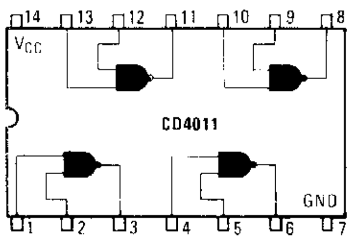
Fig. 94-2

An NE555 timer is used to generate an ultrasonic signal in the 20- to 65-kHz range. The speaker is a small piezoelectric tweeter with response above 20 kHz. These frequencies are said to be annoying to rats, mice, and insects.

ULTRASONIC PEST REPELLER II



A



B

RADIO-ELECTRONICS

Fig. 94-3

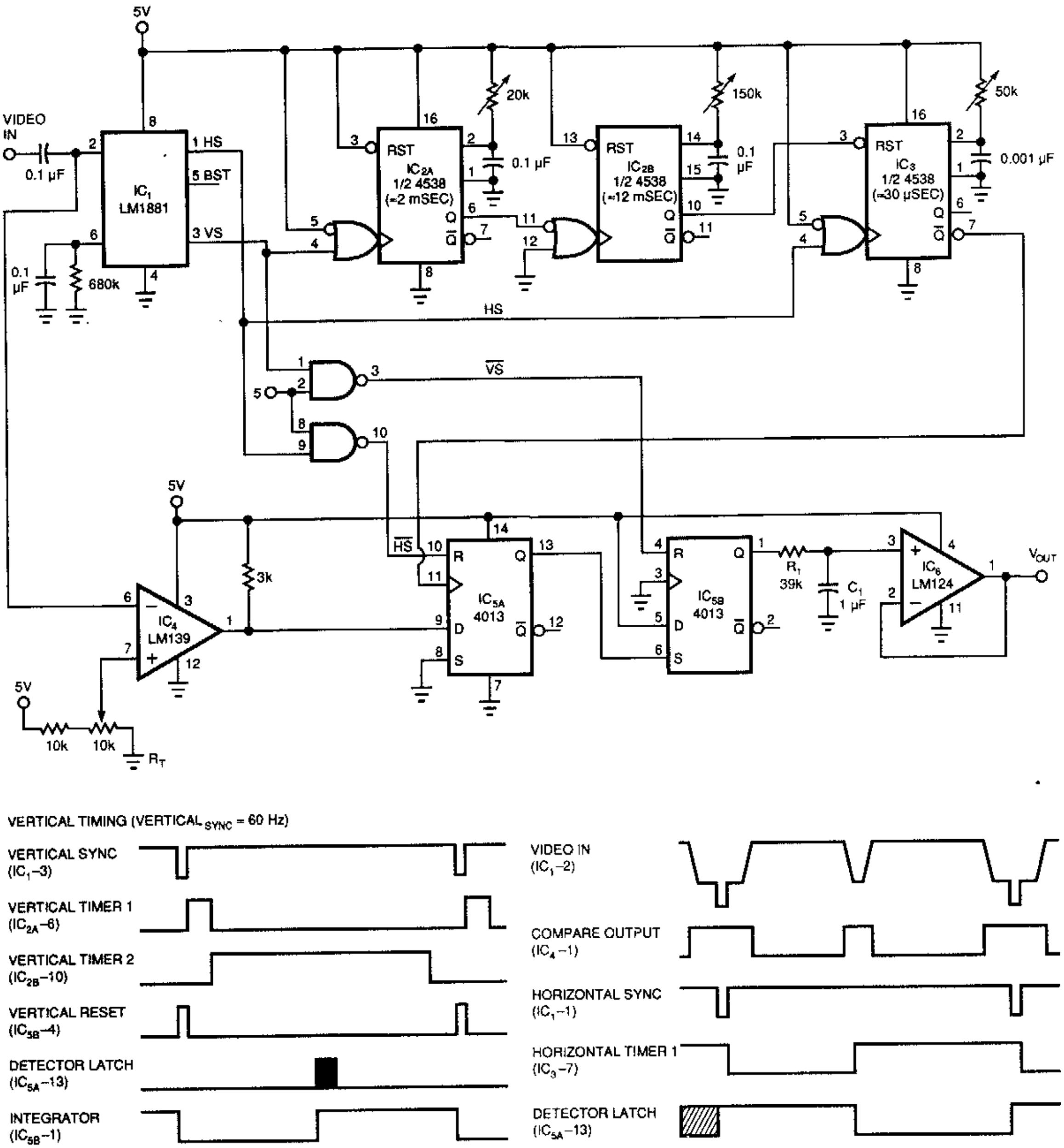
A CD4011 Quad NAND gate acts as an oscillator, operating around 40 kHz. The small amount of filtering used modulates this with 120-Hz hum. The speaker is a small tweeter for hi-fi applications.

95

Video Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 680. The figure number in the box of each circuit correlates to the entry in the Sources section.

ANALOG VOLTAGE CAMERA-IMAGE TRACKER



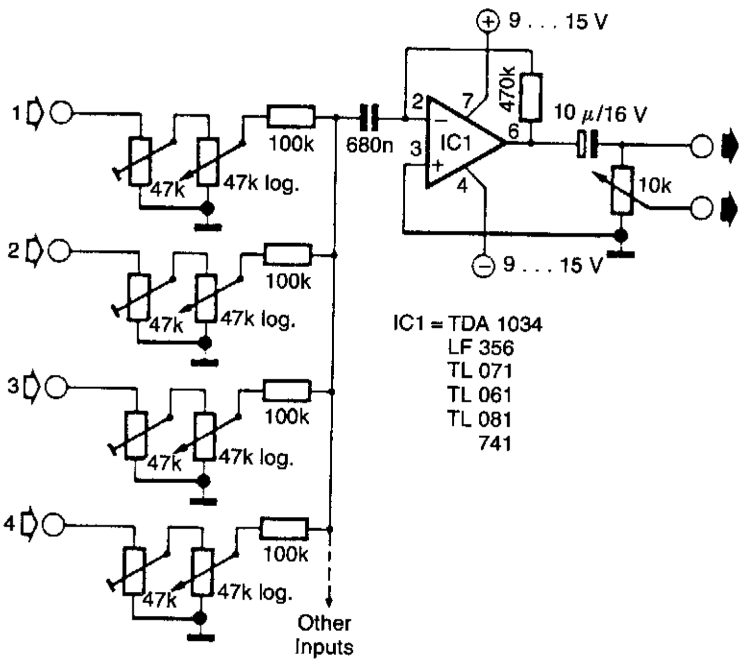
NOTE: ASSUME TARGET IS IN MIDDLE OF SCREEN

EDN

Fig. 95-1

By using a low-cost RS-170 camera and this circuit, a voltage that trades the position of an object in the field of view of a camera is generated. IC2A and IC2B form a valid video gate that holds IC3 in reset during the internal vertical blanking to prevent false interpretation of the UBI as black video. IC4 is a black level detector. The circuit tests for a black object in the middle of each video line. IC5 latches the comparator's output and produces a square wave whose duty cycle depends on where the black level is detected in the video field. R1, C1, and IC6 integrate and buffer the analog output voltage.

HIGH-PERFORMANCE VIDEO MIXER



ELEKTOR ELECTRONICS

Fig. 95-2

This circuit mixes H synch, V synch, and actual video. T2 mixes the synch, while T1 serves as an emitter-follower. Bandwidths of up to 25 MHz are typical for this circuit.

VIDEO A/D—D/A CONVERTER

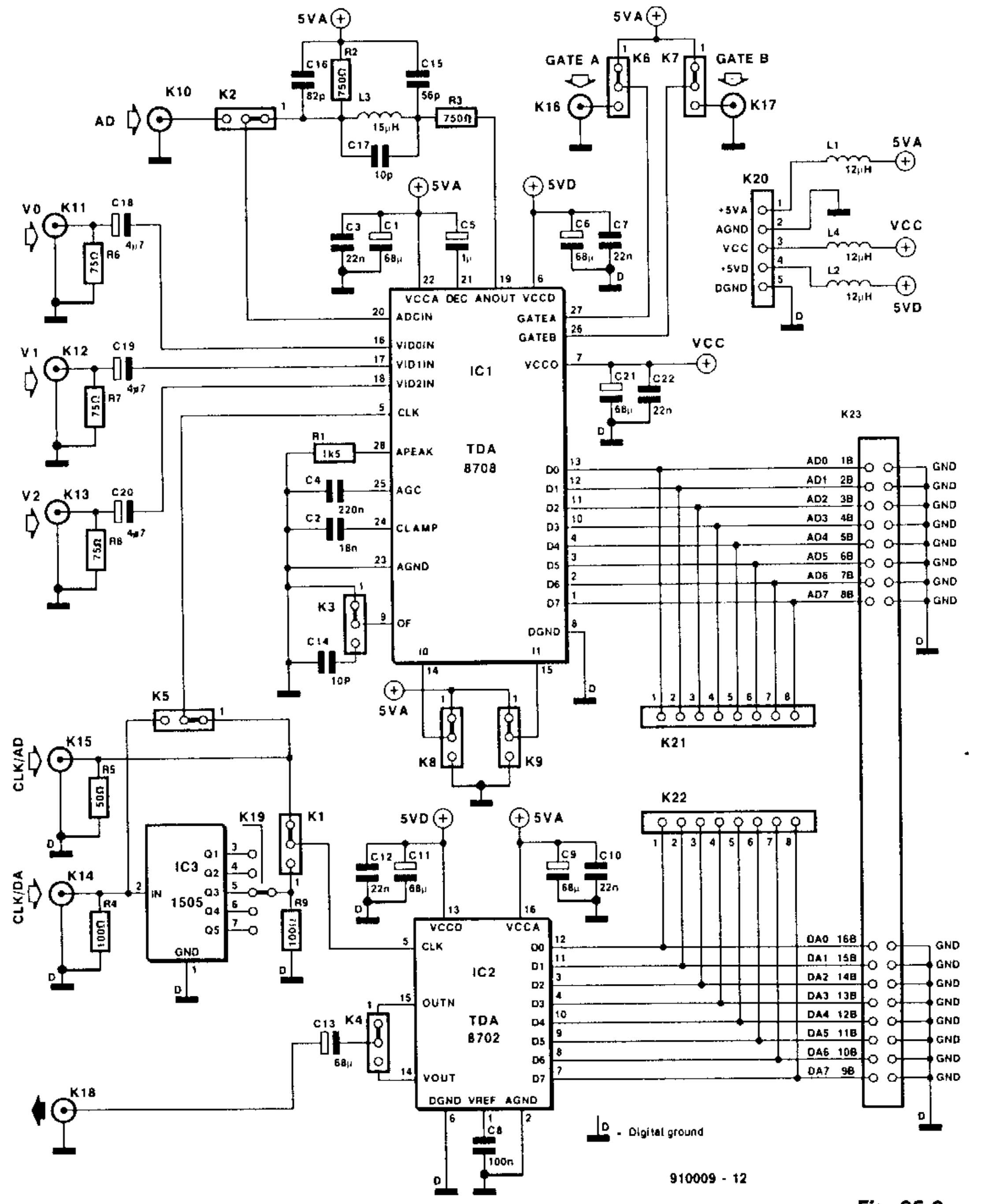
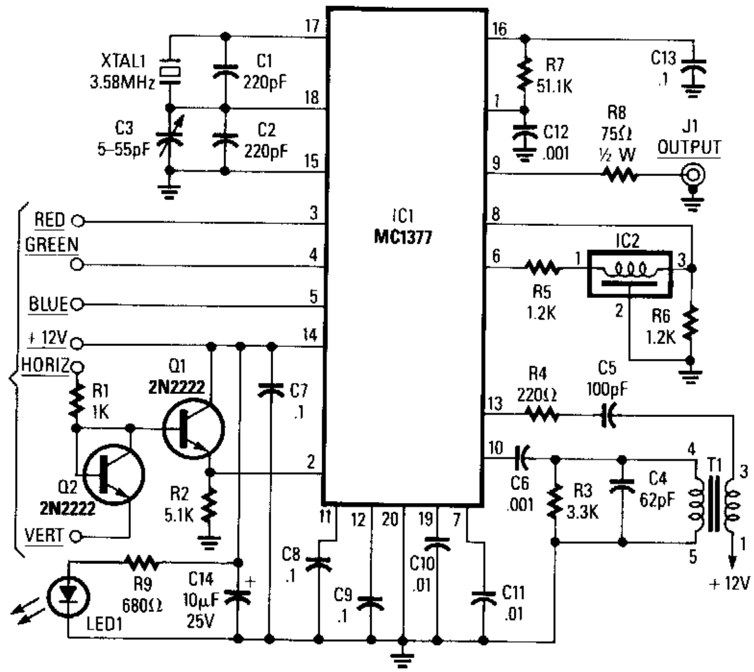


Fig. 95-3

VIDEO A/D—D/A CONVERTER *(Cont.)*

This circuit is useful for digital video experiments and for interfacing video with a computer that has a TDA8708 (Philips). The A/D converter provides 8-bit digitized video to k23 socket and k21 socket. A TDA8702 (Philips) D/A converter recovers analog video. IC3 is a 1- to 5-ms delay line (1505) to delay clock pulses in 1-ms steps to the D/A converter. Clock speeds can be up to 30 MHz. Three video inputs are provided for three analog channels (e.g., R, G, B video).

RGB/NTSC CONVERTER

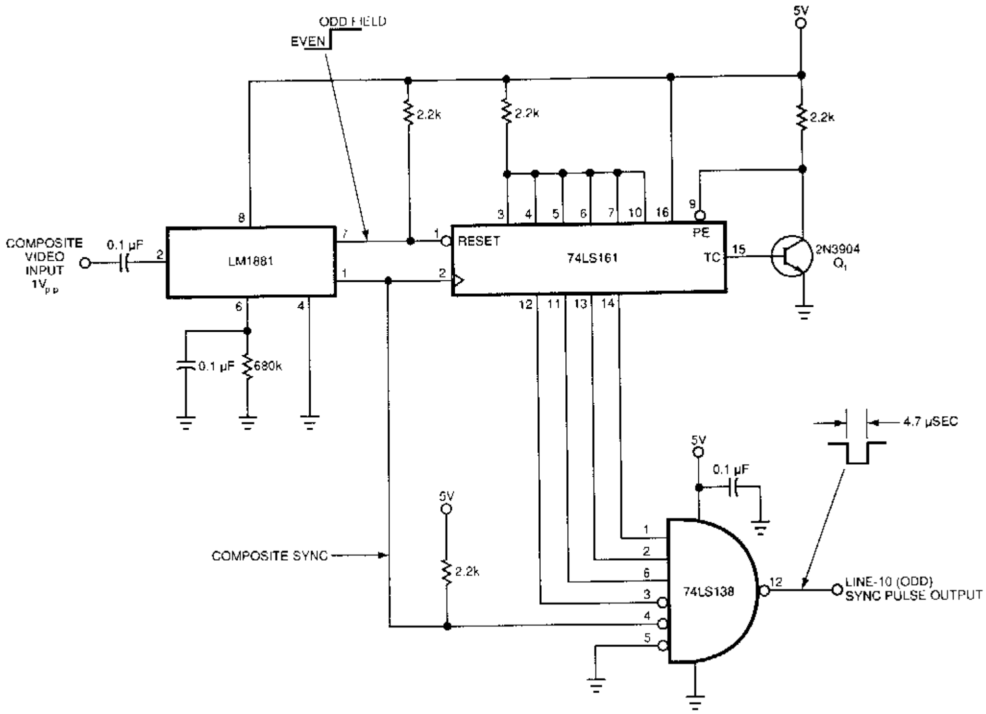


RADIO-ELECTRONICS

Fig. 95-4

Using a Motorola MC1377, this circuit produces NTSC video from an RGB source. Components are not critical, except for R7 and CB, which should be 1% and 2% tolerance, respectively.

TV LINE PULSE EXTRACTOR



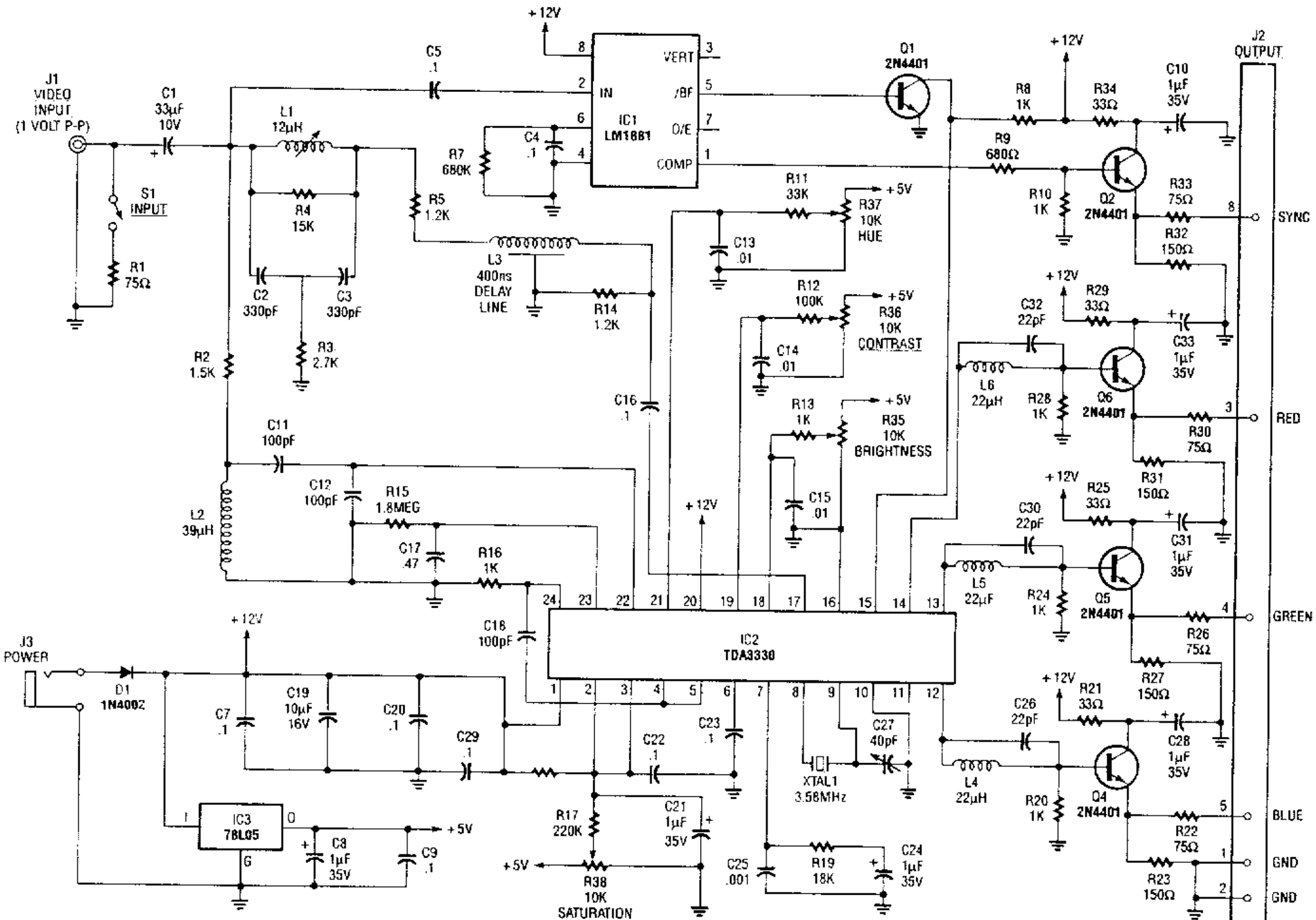
EDN

Fig. 95-5

This circuit uses a sync-generator chip, a counter, and a decoder to detect the horizontal sync pulse that occurs at the beginning of line 10 in field 1 of an NTSC television picture. You can use this circuit to compare the time delay between sync signals at various locations, and to determine and correct for any drift between the two master clocks.

The output of the LM1881 sync separator is the key to detecting line 10; the odd/even line goes high on the leading edge of the first equalizing pulse in the middle of line 4. Thus, you can use this knowledge to find virtually any other line in the field. This particular circuit locates line 10 of field one. The circuit resets the 74LS161 counter until the odd/even line goes high. Then, 74LS161 counts the positive transitions of the sync signal. After 11 positive transitions, the sync pulse drives pin 4 of the 74LS138 decoder low, and the line-10 sync pulse appears at pin 12 of the decoder. (The circuit counts to 11, as opposed to the 6 you might expect—because the composite sync signal contains more than 1 pulse per line). The counter remains in its maximum-count state until the sync separator causes a reset because Q1 feeds the inverted terminal-count output back to the parallel-enable input.

NTSC/RGB VIDEO DECODER

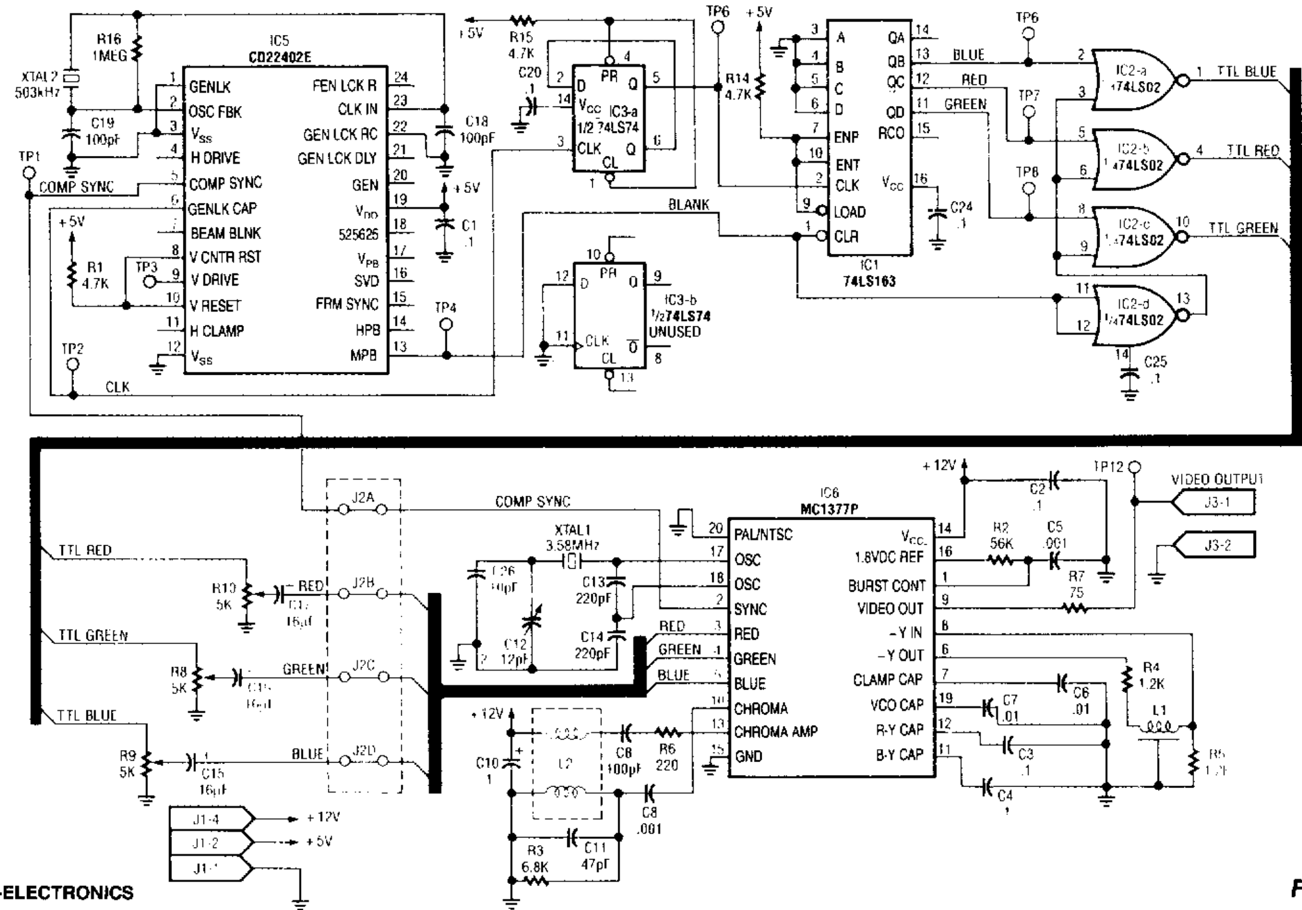


RADIO-ELECTRONICS

Fig. 95-6

An NTSC/RGB decoder is shown here. Using a TDA3330, 1-V input video is broken down into its R, G, B components, and composite synch. U1 is an integrated synch separator (LM1881). This circuit should be useful for interfacing RGB monitors to NTSC video systems.

COLOR-BAR GENERATOR

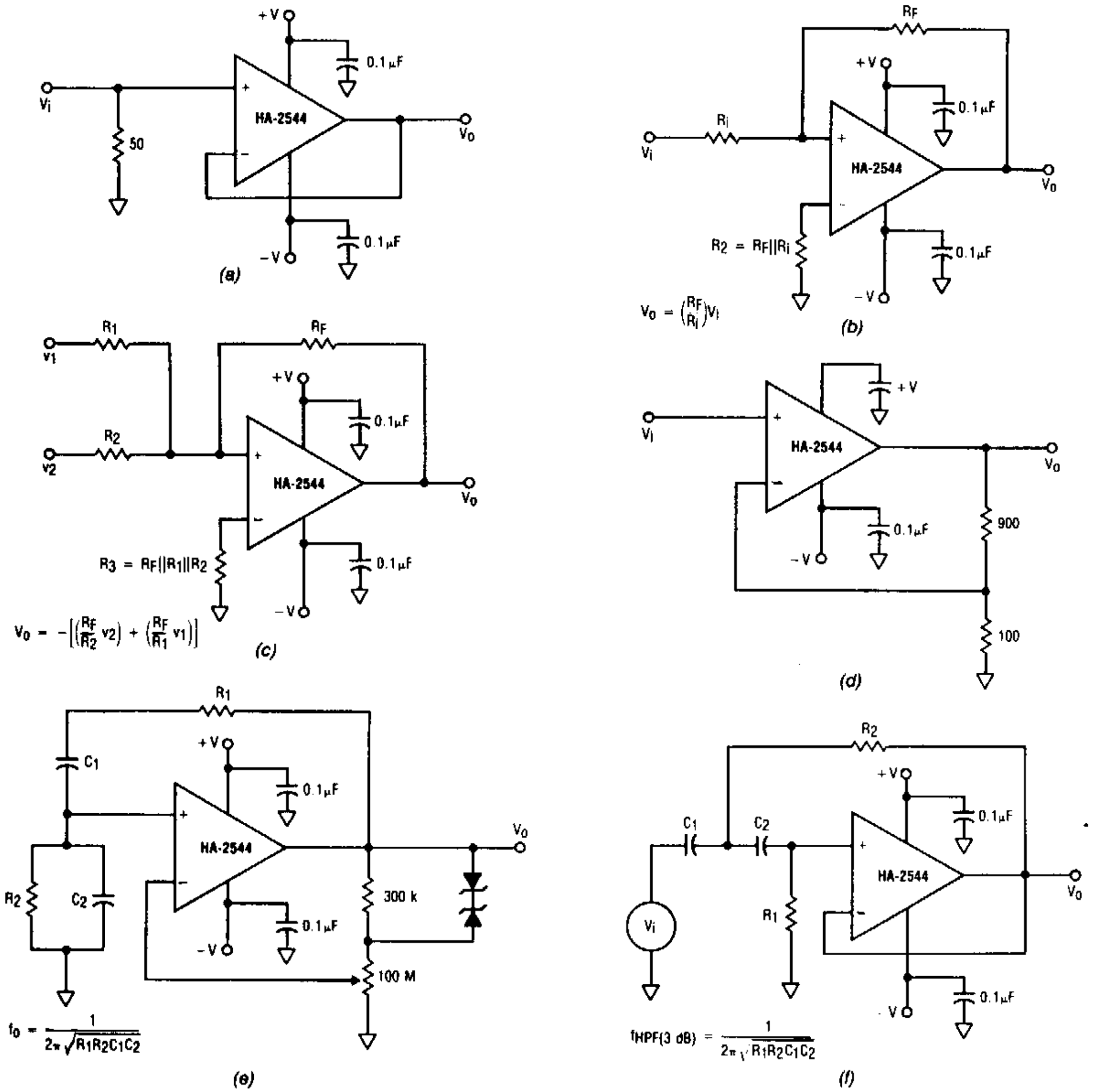


RADIO-ELECTRONICS

Fig. 95-7

IC5 generates RS-170 NTSC synch. IC1, IC2, and IC3 make up the red, green, and blue video signals that drive the video encoder section of IC6 to make up the color bars. IC3 is an $a \div 2$ counter that drives 4-bit counter IC1. Gates IC2a through IC2b form the R, B, and G, video signals. IC6 encodes these, plus synch, to form an NTSC video output signal, which appears at TP12.

VIDEO OP AMP CIRCUITS

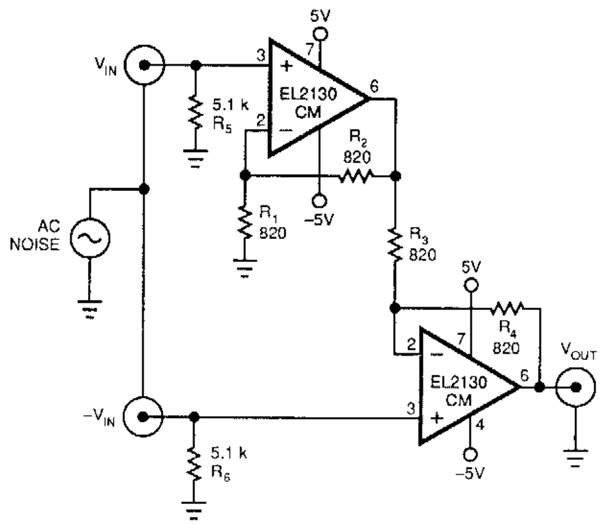


ELECTRONIC PRODUCTS ***Fig. 95-8***

These 6 circuits use the Harris HA2544 op amp. Component values are obtained by using the equations in the figure.

The HA-2544 can be used in any number of standard op amp configurations, including a voltage follower (a), an inverting amplifier (b), an inverting summer (c), a buffer amplifier with a gain of 10 (d), a Wien-bridge oscillator with zener diode adaptive feedback (e), and a second-order, high-pass active filter (f).

VIDEO LOOP-THRU AMPLIFIER



This video bandwidth amplifier rejects common-mode noise, such as 60- and 120-Hz hum. Bandwidth is typically 60 MHz for a differential gain of 2. The common-mode rejection is typically 45 dB.

The design equations are:

$$A_{DIFF} \geq 1,$$

$$R_2 = R_4 \text{ (select both for optimum bandwidth)}$$

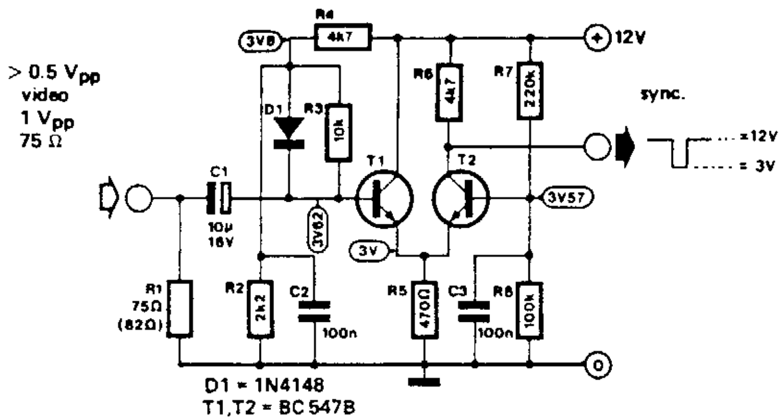
$$R_1 = R_2\,(A_{DIFF} - 1)$$

$$R_3 = R_4\,(A_{DIFF} - 1)$$

EDN

Fig. 95-9

SYNC SEPARATOR

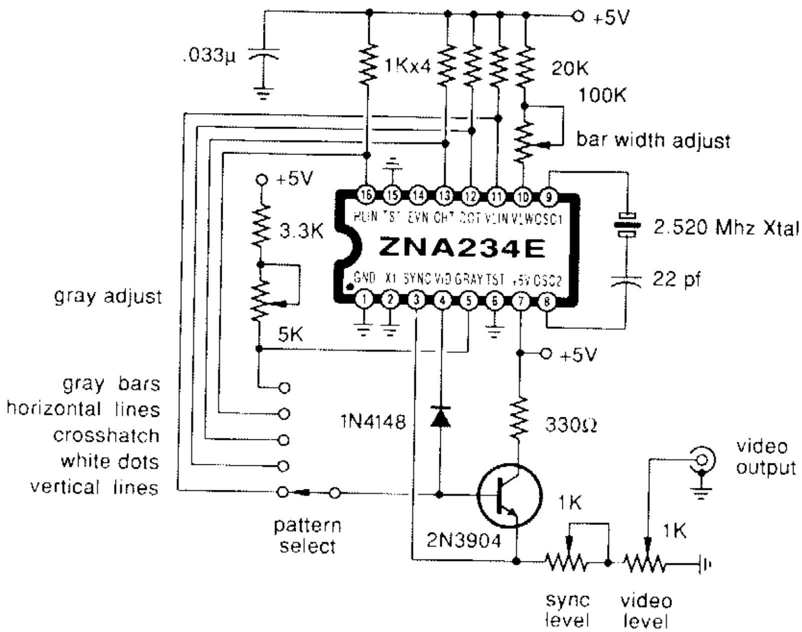


ELEKTOR ELECTRONICS

Fig. 95-10

This circuit separates the synch from the composite video signal. It uses a two-transistor comparator. Output is 9 Vpp with a 0.5-Vpp (minimum) video input signal.

SIMPLE MONOCHROME TV-PATTERN GENERATOR



RADIO-ELECTRONICS

Fig. 95-11

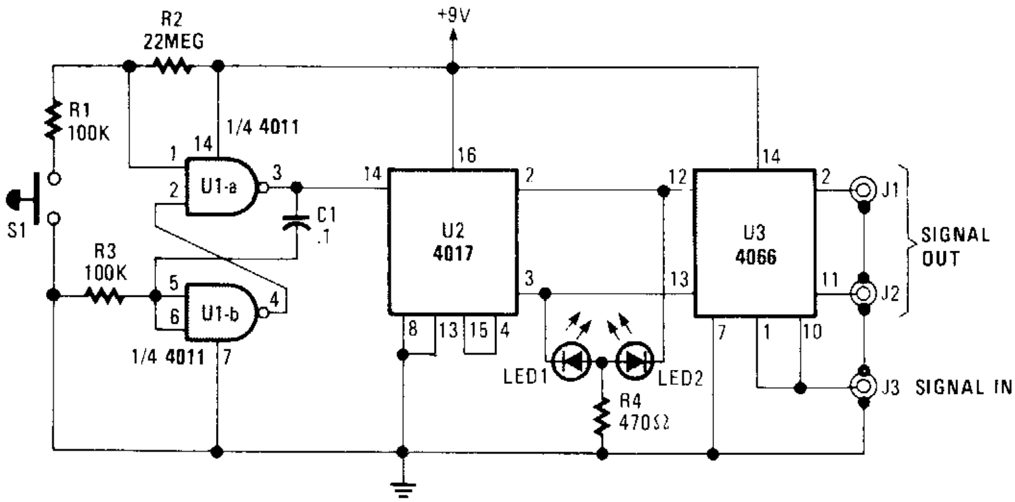
Using a Plessey ZNA234E IC, this generator produces sync, blanking, gray bars, lines, dots, and crosshatch patterns.

96

Video-Switching Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 680. The figure number in the box of each circuit correlates to the entry in the Sources section.

REMOTE SELECTION VIDEO SWITCH

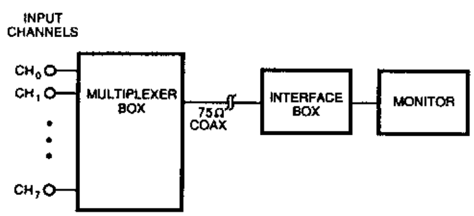


POPULAR ELECTRONICS *Fig. 96-1*

The A/B Switch circuit consists of three ICs and a handful of resistors. Two gates from a 4011 quad 2-input NAND gate (U1A and U1B) are configured as a monostable multivibrator that, when switch S1 is pressed, triggers a 4017 decade counter/divider, which has been set to recycle after a count of two. The outputs of U2 at pins 2 and 3 are fed to the control inputs of U3 (a 4066 quad bilateral switch) at pins 12 and 13. Depending on which control input is high, either the J1 or J2 output is selected.

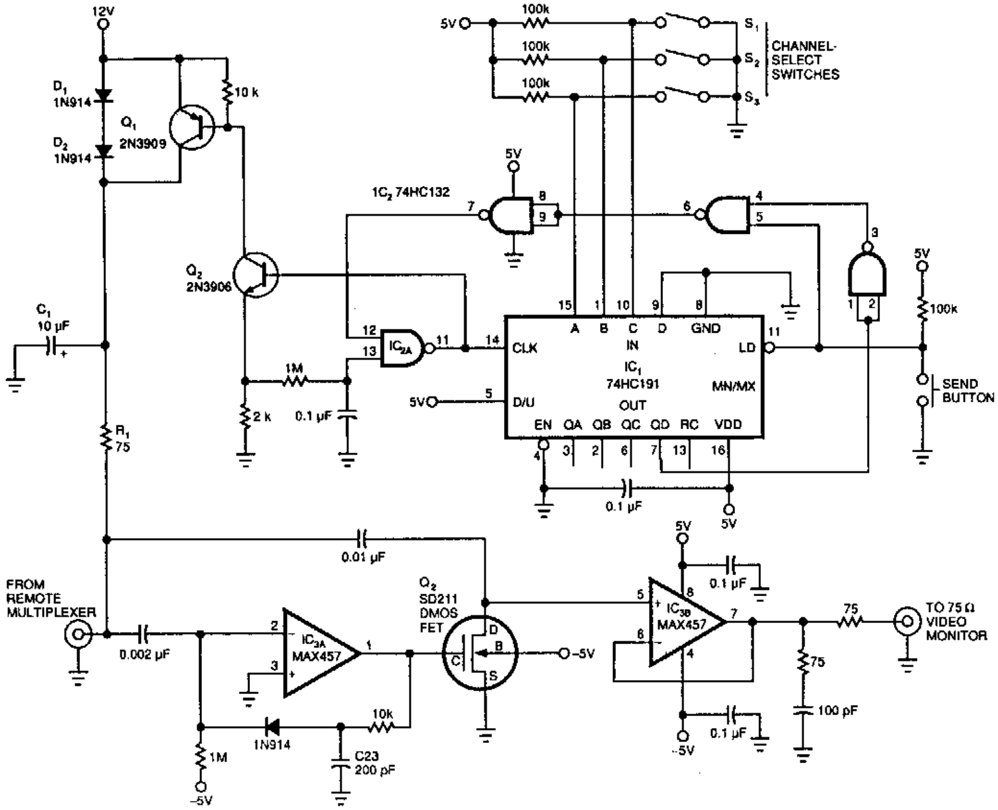
With a little modification, the switch could be set to trigger at a set rate (automatically). With the addition of another 4066, it could have as many as 8 channels. One possible application would be in a security surveillance system.

REMOTE-CONTROLLED SWITCHER



This 1-cable system carries composite video (NTSC, PAL, or SECAM), power, and channel-select signals.

EDN *Fig. 96-2(a)*

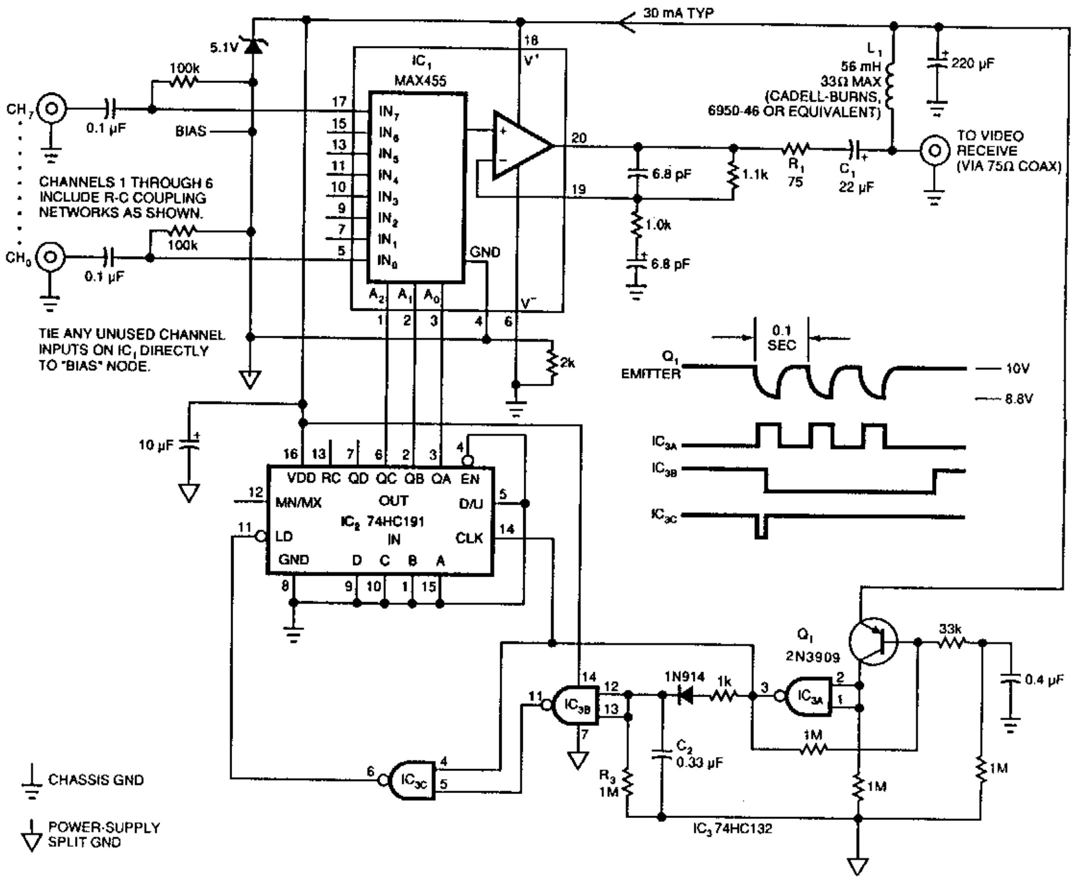


EDN

Fig. 96-2(b)

The interface end of Fig. 96-2(a)'s circuit deliver 10 V down the cable, pulses the supply voltage to transmit channel-change commands, and buffers the received video signal.

The multiplexer circuit in Fig. 96-2(a) receives power and control signals over the coaxial cable, while driving the cable with the currently selected video signal.



EDN

Fig. 96-2(c)

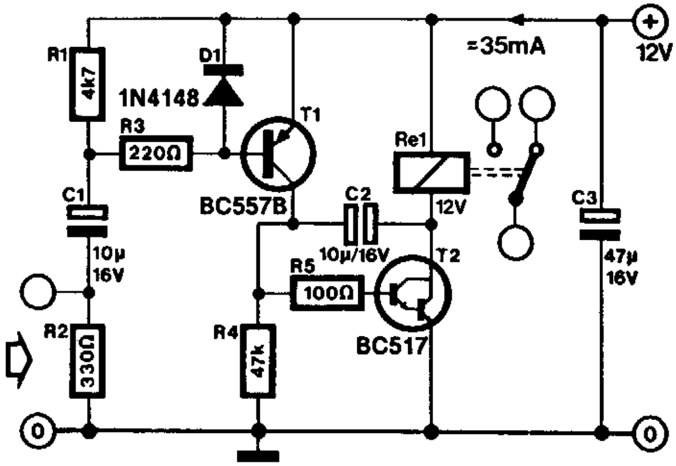
The interface circuit (Fig. 96-2(b)) delivers 10 V to the cable and pulses the supply voltage to select one of 8 channels. When the send button is depressed, a digital burst of 1.2 V amplitude (negative) is superimposed on the 12-V line (as a voltage drop). This does not affect the video signal. The multiplexer circuit (Fig. 96-2(c)) consists of a multiplexer and an amplifier. The multiplexer is a Maxim MAX455. The digital code on the supply line is picked off by A1, IC3A, and interfaced to counter IC2, which drives the multiplexer to select the desired video channel.

97

Voice-Operated Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 681. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE VOX



ELEKTOR ELECTRONICS

Fig. 97-1

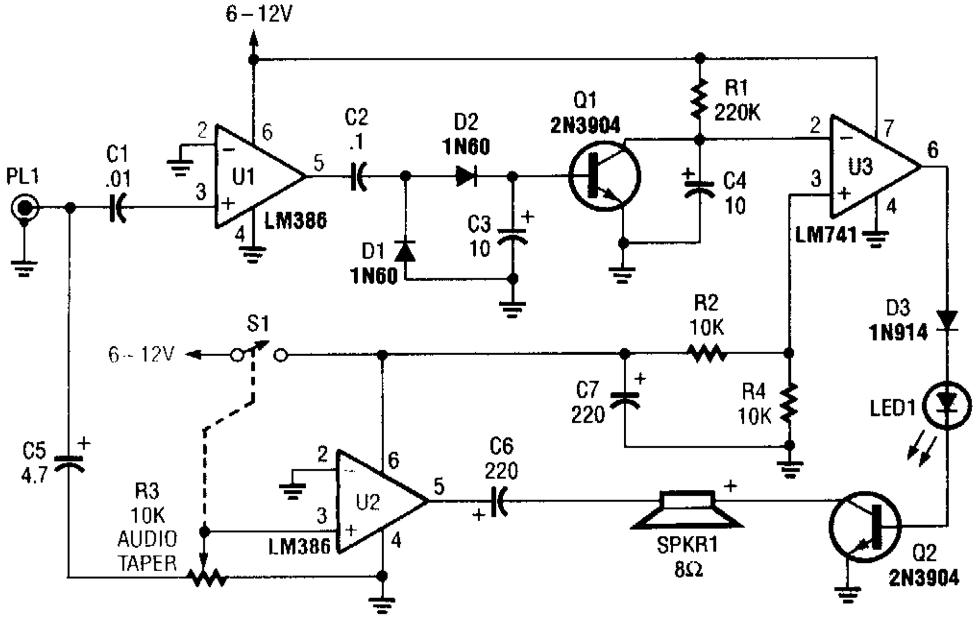
A *VOX* is a voice-operated switch that is often used as a substitute for the press-to-talk switch on a microphone. This VOX can be connected to almost any audio equipment that has a socket for an external loudspeaker. The actuation threshold is set by the volume control on the AF amplifier that drives the VOX.

The (loudspeaker) signal across R2 is capacitively fed to the base of T1. Resistor R3 limits the base current of this transistor when the input voltage exceeds 600 mV. Diode D1 blocks the positive excursions of the input signal, so that V_{eb} cannot become more negative than about 0.6 V.

The output relay is driven by Darlington T2. Resistor R4 keeps the relay disabled when T1 is off. The value of bipolar capacitor C2 allows it to serve as a ripple filter in conjunction with T2. Resistor R5 limits the base current of T2 to a safe level.

The switching threshold of the VOX is about 600 mV across R2. The maximum input voltage is determined by the maximum permissible dissipation of R2 and R3. As a general rule, the input voltage should not exceed 40 Vpp. The current drawn by the VOX is mainly the sum of the currents through the relay coil and through R5. The resistor can carry up to 100 mA when the VOX is overdriven.

SCANNER VOICE SQUELCH



POPULAR ELECTRONICS

Fig. 97-2

This circuit detects the presence of audio (voice) on the output of a scanner. If the scanner stops on a "dead carrier" or noise, the circuit mutes the speaker to avoid annoying noise.

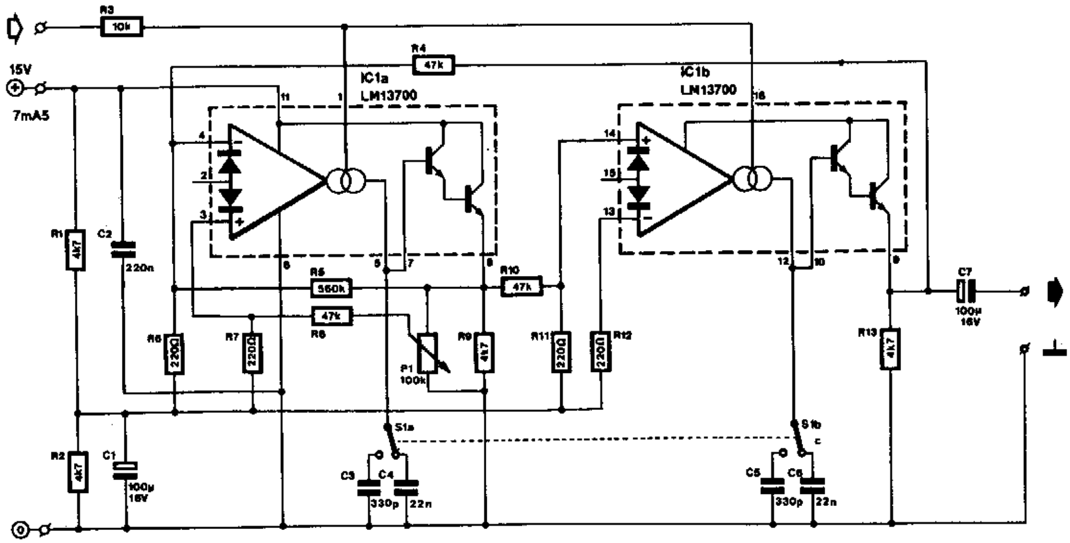
U1 amplifies speech and drives rectifier D1/D2 and switch Q1. Comparator U3 drives speaker switch Q1 and indicator LED1. Q2 completes the speaker path to ground. U2 is an audio amplifier to drive the speaker. R3 is a volume control. PL1 connects to the scanner speaker or to the headphone jack.

98

Voltage-Controlled Oscillators

The sources of the following circuits are contained in the Sources section, which begins on page 681. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE AUDIO-FREQUENCY VCO



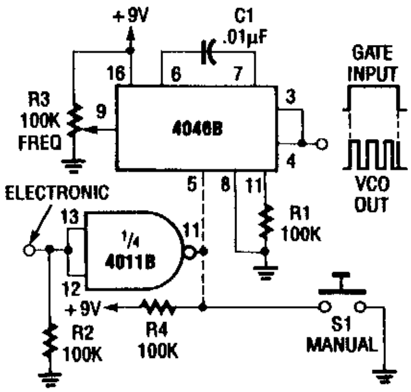
ELEKTOR ELECTRONICS

Fig. 98-1

The frequency of this sine-wave oscillator is determined by a direct voltage, U_c, of 0 to 15 V. The distortion on output signals of up to 10 Vpp is not greater than 1%. When the output is reduced with the aid of P1 to 1 Vpp, the distortion drops to below 0.1%. It is not recommended to use output signals below 1 Vpp, because the oscillator then becomes unstable and temperature-dependent.

The oscillator consists of two operational transconductance amplifiers (OTAs) contained in one package. Their Amp-bias inputs, pins 1 and 16, are connected in parallel. These inputs can drive the output currents at pins 5 and 12 to a peak value of up to 0.75 mA.

Switch S1 enables the oscillator output to be set to two ranges: 6.7 to 400 Hz and 400 Hz to 23.8 kHz. The overall range needs a control voltage that varies from 1.34 to 15 V. When the frequency is changed by a variation of U_C and the setting of P1 is not altered, the output signal might be distorted. In other words, the amplitude of the signal must be adapted to the frequency.

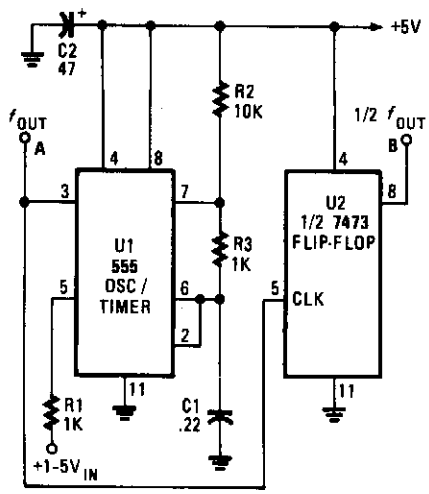


GATED WIDE-RANGE VCO

A CD4046 can be gated either with a switch or electronically, as shown in the figure. Frequency range of this circuit is to 1.5 kHz; use another C_1 for higher frequencies.

RADIO-ELECTRONICS

Fig. 98-2

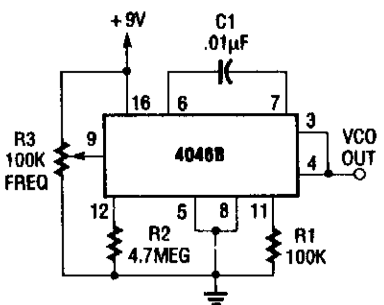


SIMPLE 555 VCO

The VCO has an output frequency that ranges from 1 500 Hz at $V_{in} = 1$ V to 300 Hz at $V_{in} = 5$ V. R_1 or C_1 can be varied to change this range. U2 provides a symmetrical square-wave output of half the timer frequency.

POPULAR ELECTRONICS

Fig. 98-3



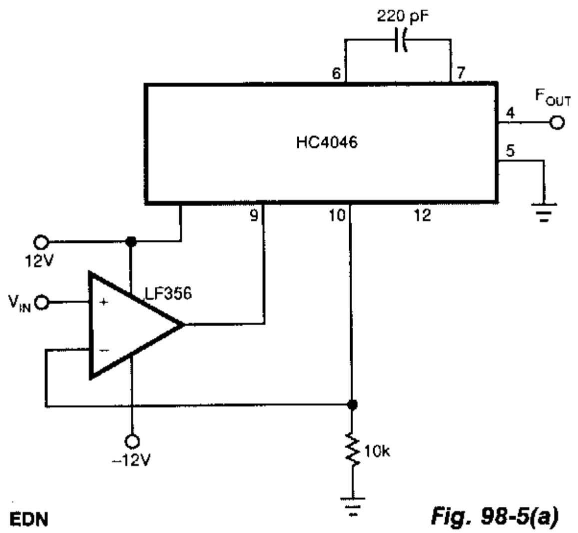
RESTRICTED-RANGE VCO

This VCO is adjustable from 60 Hz to 1.4 kHz. C_1 can be changed for other ranges.

RADIO-ELECTRONICS

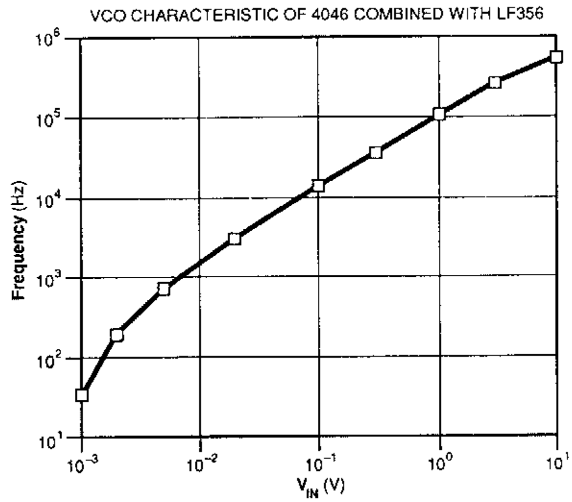
Fig. 98-4

LINEAR VCO



EDN

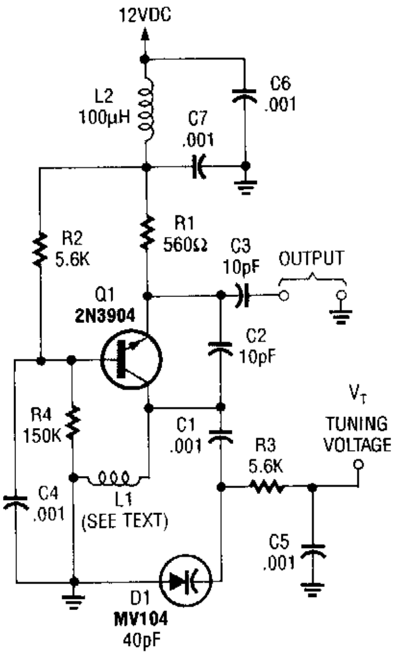
Fig. 98-5(a)



EDN

Fig. 98-5(b)

This VCO uses an LF356 op amp to produce a linear frequency vs. voltage characteristic using the CMOS HC4046. The frequency range can be changed by changing the capacitor connected between pins 6 and 7 of the HC4046. Using the HC4046's internal transistor instead of an external component achieves the linearization in Fig. 98-5(b).



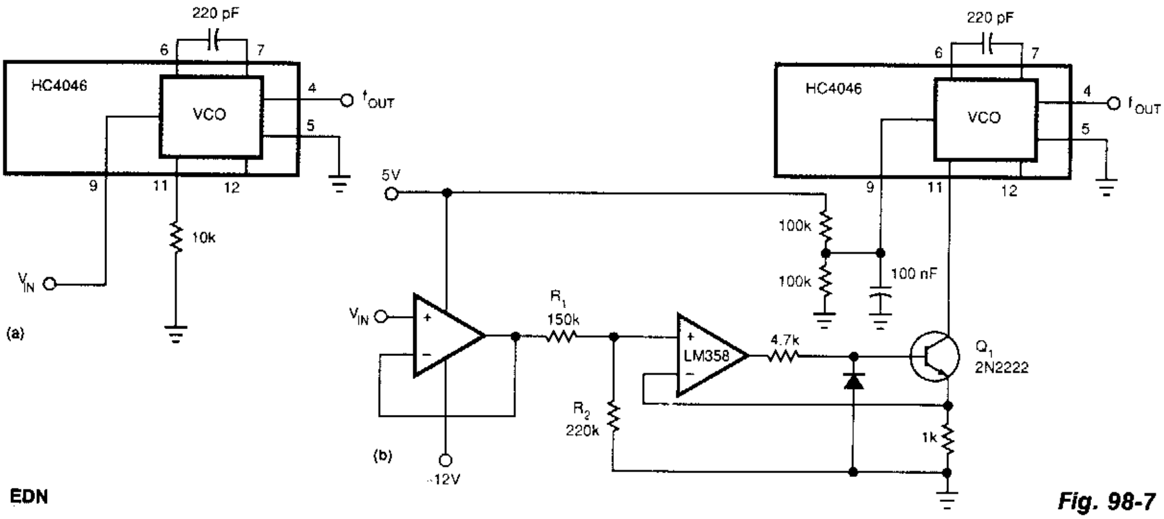
POPULAR ELECTRONICS

Fig. 98-6

VOLTAGE-TUNED VHF OSCILLATOR

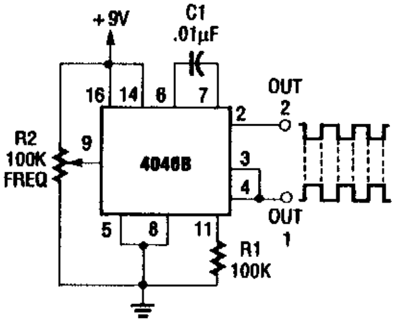
This VHF VCO circuit is suitable for 30 to 200 MHz. Q1 can be replaced by a 2N3563 for operation above 100 MHz. L1 is chosen to resonate to the desired frequency with the varactor capacitance of 40 pF. Other varactors can be substituted or two back-to-back varactors can be used for better linearity, depending on the application.

VOLTAGE-CONTROLLED CURRENT SINK



EDN *Fig. 98-7*

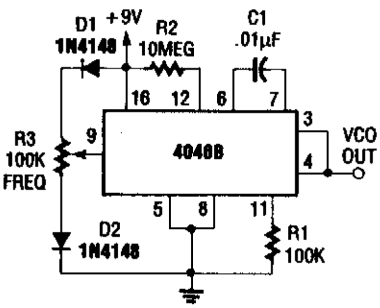
This circuit widens the linear frequency span of an HC4046 from one decade to nearly three decades. An LM358 is used as a constant-current sink to replace the frequency-determining resistor (10 kΩ) from pin 9 to ground. Pin 9 is held at a fixed 2.5 V for this application.



RADIO-ELECTRONICS *Fig. 98-8*

BIPHASE WIDE-RANGE VCO

Using a CD4046B, this circuit generates a biphase signal. The frequency range is below 100 Hz to about 1.5 kHz.

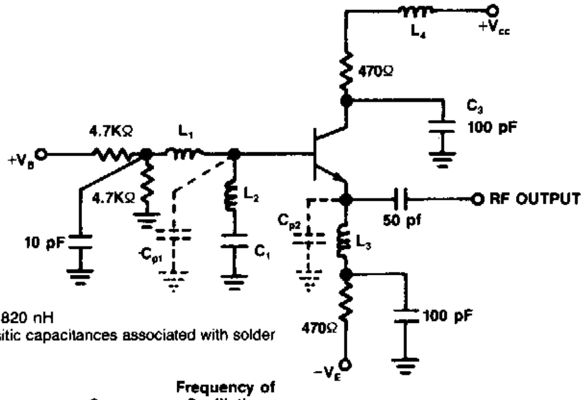


RADIO-ELECTRONICS *Fig. 98-9*

WIDE-RANGE VCO

This circuit covers 0 to 1.4 kHz. C_1 can be changed to cover other ranges, as desired.

VARACTORLESS VCO



- L_1 and L_4 — 680 — 820 nH
- C_{p1} and C_{p2} are parasitic capacitances associated with solder pads.

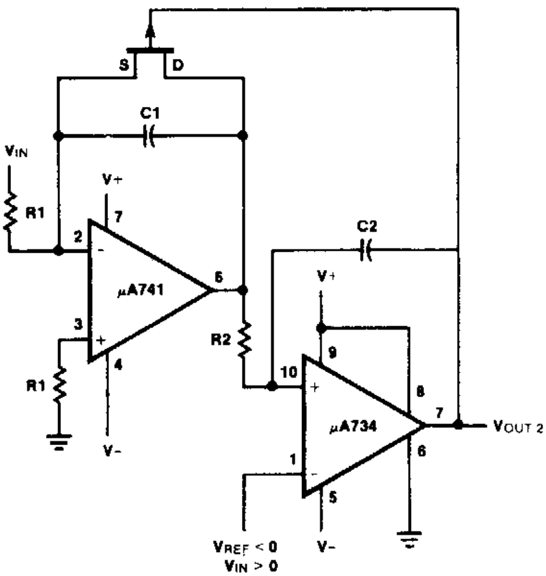
L_2, L_3	Transistor	C_1	Frequency of Oscillation
120 nH	2N4124	470 pF	310 MHz
70 nH	MRF 901	470 pF	700 MHz
50 nH	MRF 901	470 pF	800 MHz

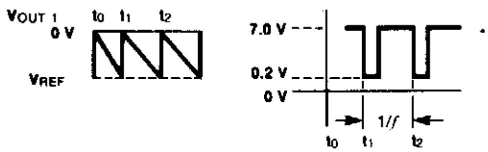
RF DESIGN

Fig. 98-10

The varactorless VCO utilizes a modified Clapp-oscillator configuration, together with some data on the type of transistor, circuit values, and operating frequencies.

VCO





$$f = \frac{V_{IN}}{|V_{REF}|R_1C_1} \qquad R_2C_2 > \frac{|V_{REF}|C_1}{I_{DSS}}$$

FAIRCHILD CAMERA

Fig. 98-11

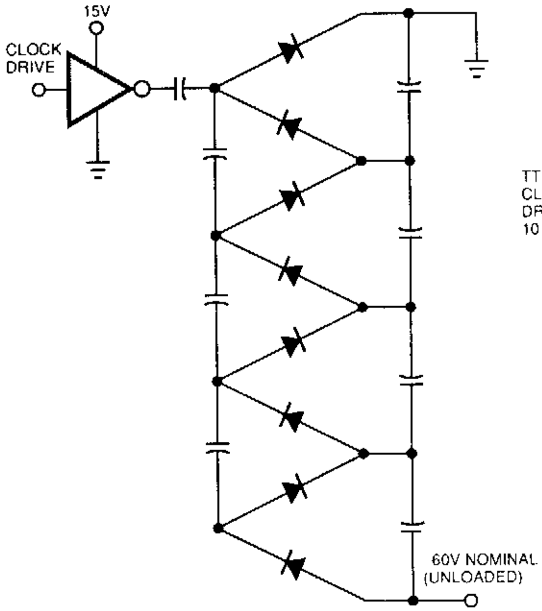
Q1, an FET, is used as a variable resistance to control frequency of oscillator.

99

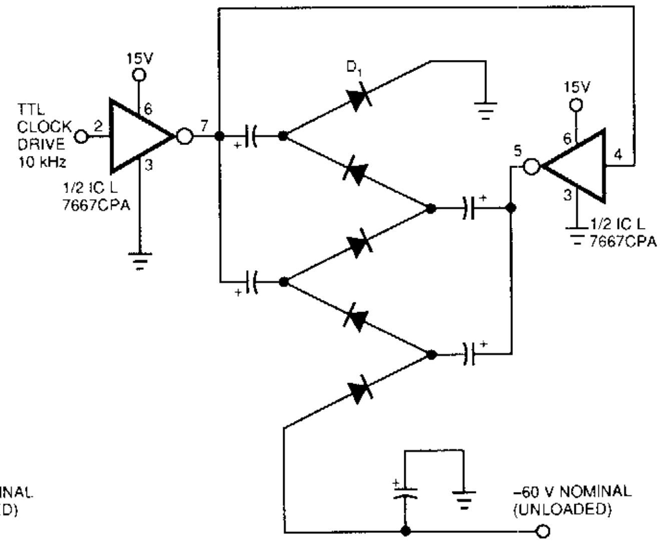
Voltage Multiplier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 681. The figure number in the box of each circuit correlates to the entry in the Sources section.

VOLTAGE MULTIPLIER



EDN Fig. 99-1(a)



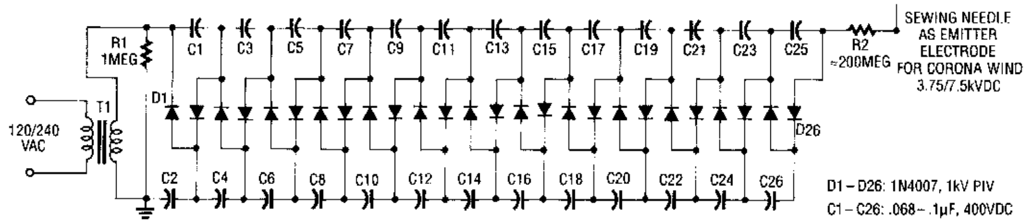
EDN Fig. 99-1(b)

Figure 99-1(a)'s circuit exhibits a high-output impedance as a result of the small effective capacitance of the series-connected capacitors, and it exhibits considerable voltage loss due to all of the diode drops. Further, this circuit requires $2n$ diodes and $2n$ capacitors to produce a dc output voltage approximately n times the rail voltage.

Figure 99-1(b)'s circuit multiplies more effectively using fewer diodes and capacitors. The parallel arrangement of the capacitors lets you use smaller capacitors than those required in Fig. 99-1(a). Alternatively, when using the same capacitor values of Fig. 99-1(a), the output impedance will be lower.

Whereas the clock source directly drives only one of the two strings of capacitors in Fig. 99-1(a), Fig. 99-1(b)'s clock drives both strings with opposite phases. This drive scheme doubles the voltage per stage of two diodes. A final diode is necessary to pick off the dc output voltage because both strings of capacitors now carry the p – p ac input-voltage waveform. The ICL7667 dual-FET driver accepts a TTL drive swing and provides a low-impedance push-pull drive to the diode string. This low impedance is particularly helpful when using a long string to raise output voltage to more than 100 V, starting from a low rail voltage.

CORONA WIND GENERATOR

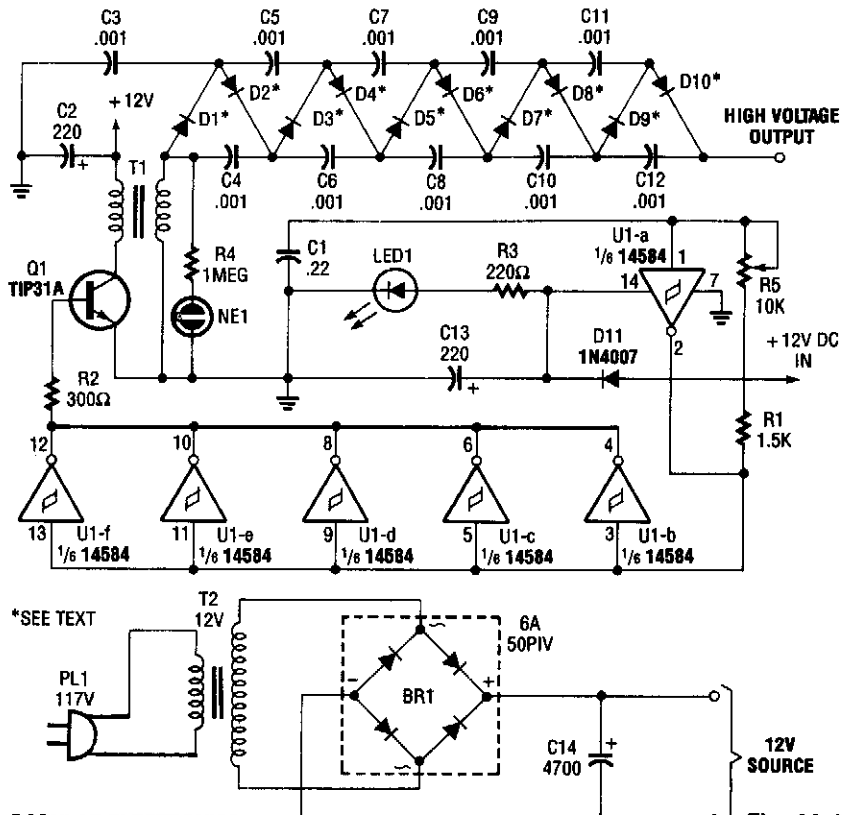


RADIO-ELECTRONICS

Fig. 99-2

This 25-stage voltage doubler will generate "corona wind." It delivers 3.75 kVdc when powered from 120 Vac, or 7.5 kVdc when powered from 240 Vac.

10 000-Vdc SUPPLY

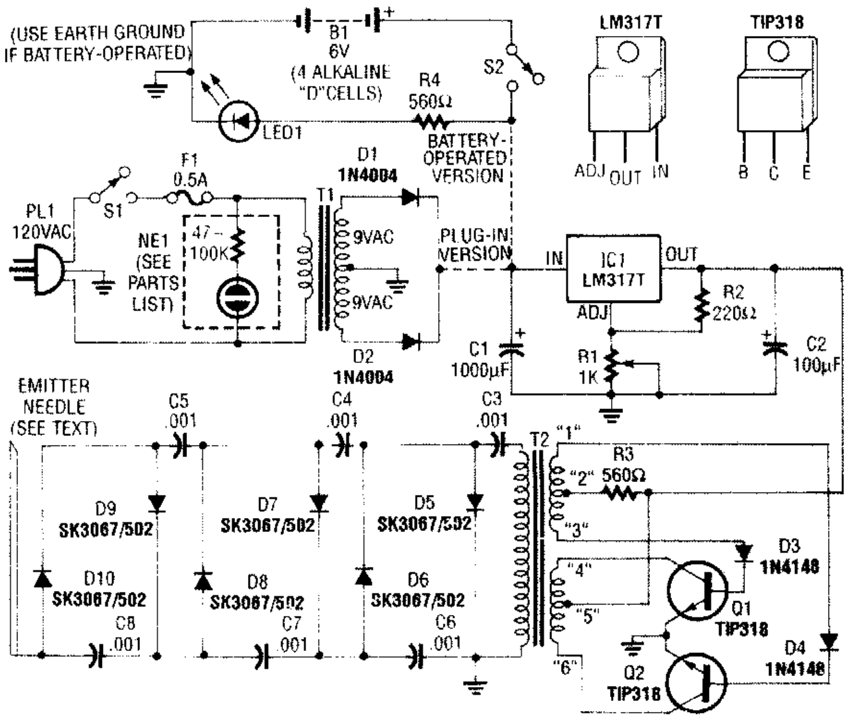


POPULAR ELECTRONICS

Fig. 99-3

A CMOS oscillator (U1A) drives. U1B through U1F, which drives Q1, which generates a 12-Vpp square wave across the primary of T1. This square wave is applied to a rectifier-multiplier circuit consisting of D1 through D10 (each is two 1N4007 diodes in series) and C3 through C12. About 10 kV is available.

HIGH-VOLTAGE NEGATIVE-ION GENERATOR



RADIO-ELECTRONICS **Fig. 99-4**

A modified B/W TV flyback transformer is used in this circuit with a voltage multiplier to produce 9- to 14-kV negative voltage. This is connected to a discharge needle to produce negative ions.

VOLTAGE DOUBLERS

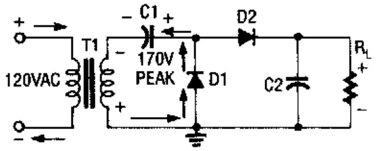


Fig. 99-5(a)

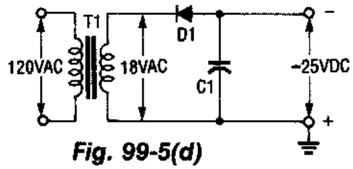


Fig. 99-5(d)

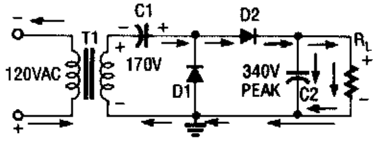


Fig.99-5(b)

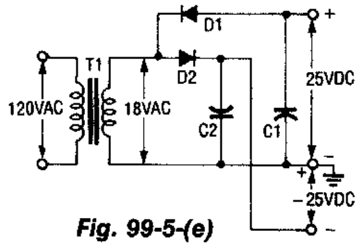


Fig. 99-5-(e)

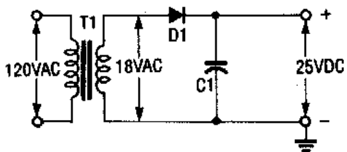
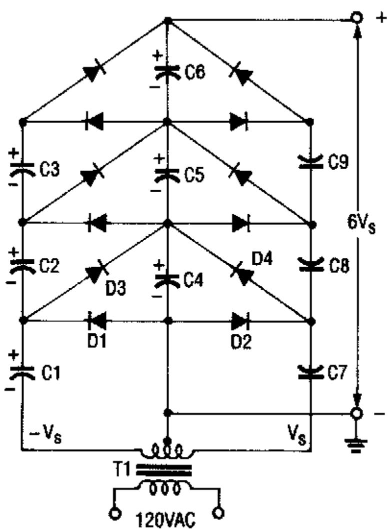


Fig. 99-5(c)

RADIO-ELECTRONICS *Fig. 99-5*

During the first half-cycle (Fig. 99-5(a)), D1 conducts, D2 cuts off, C1 charges to 170 V peak, and C2 discharges through R_L. For the second half-cycle (Fig. 99-5(b)), the input polarity is reversed, and both the input and C1 are in series, which produces 340 V (peak). Now, D1 cuts off, while D2 conducts, and the current divides between C2 and R_L; the cycle then repeats. Two Half-Wave Rectifiers, one with a positive output (Fig. 99-5(c)) and one negative (Fig. 99-5(d)), combine to make a full-wave voltage doubler (Fig. 99-5(e)).

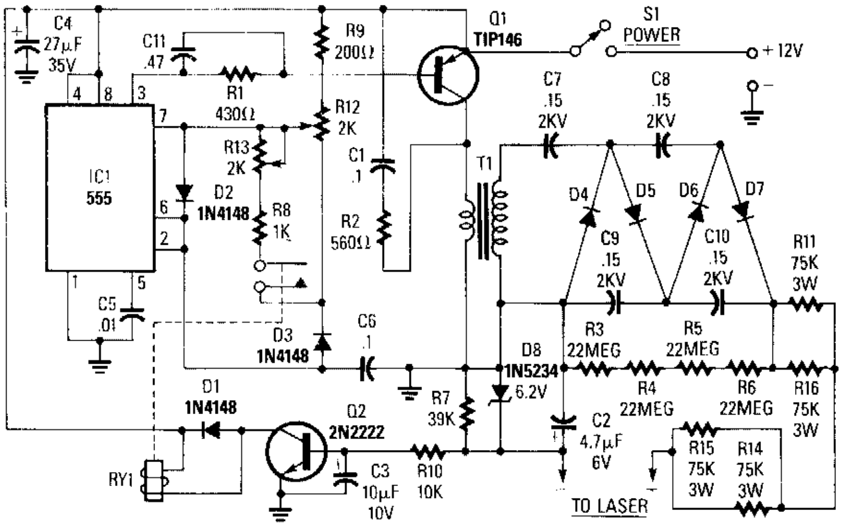


COCKCROFT-WALTON CASCADED VOLTAGE DOUBLER

A center-tapped transformer of secondary voltage. Two V_s can be used to power a voltage multiplier. For higher voltages, simply add more sections.

RADIO-ELECTRONICS *Fig. 99-6*

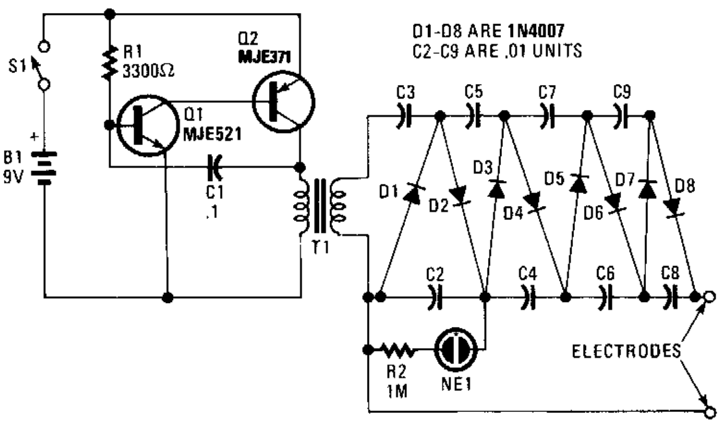
LASER POWER SUPPLY



RADIO-ELECTRONICS *Fig. 99-7*

IC1 is a 555 timer running at about 16 kHz. This IC drives Q1, a TIP146, which produces a 12-V square wave across T1 primary. This produces between 800 and 2,000 V across the secondary, which is doubled to 3 to 5 kV. When the load (laser) on the power supply increases, current Q2 is turned on, which energizes RY1. This changes the duty cycle of the 555 timer. To adjust this supply, set R12 and R13 at the center. Adjust R12 until the laser tube triggers, and make sure that the relay pulls in. If the relay chatters, adjust R12. If the full-clockwise adjustment of R12 fails to ignite the tube, adjust R13.

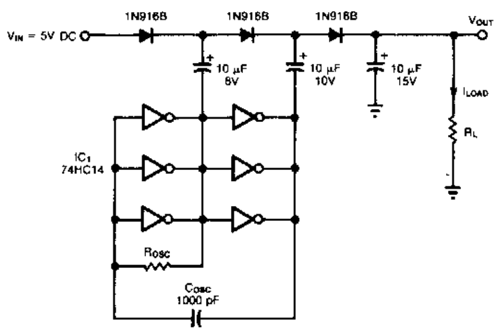
2,000-V LOW-CURRENT POWER SUPPLY



RADIO-ELECTRONICS *Fig. 99-8*

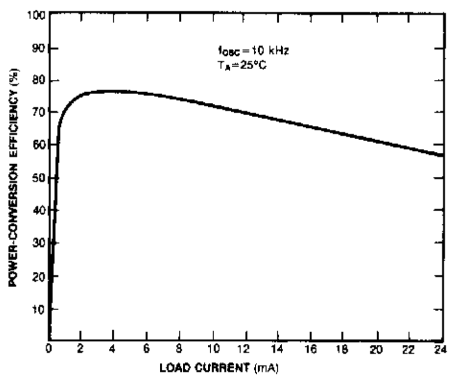
In this circuit Q1, Q2, R1, and C1 form a multivibrator. The square wave that results from the oscillation of this circuit (20 to 30 Vpp) is stepped up by T1 (an audio transformer of the type used in radios or small TVs). An 8- to 1,200-Ω impedance ratio equates to a turn ratio of 12:1. The ac from the secondary of T1 is applied to the multiplier circuit (D1 to D8 and C2 to C9). NE1/R2 are used as an operating indicator. The circuit will supply about 2,000 V. C2 to C9 should have a 400-V or higher voltage rating.

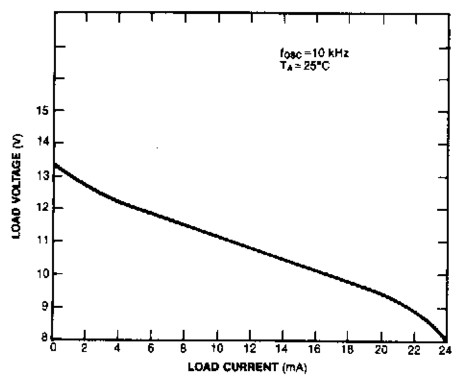
LOW-CURRENT VOLTAGE TRIPLER



This circuit generates a 15V output by tripling V_{IN}.

(a)





These curves show Fig. 99-9(a)'s power-conversion efficiency and laod voltage (b) vs load current.

(b) ***(c)***

EDN ***Fig. 99-9***

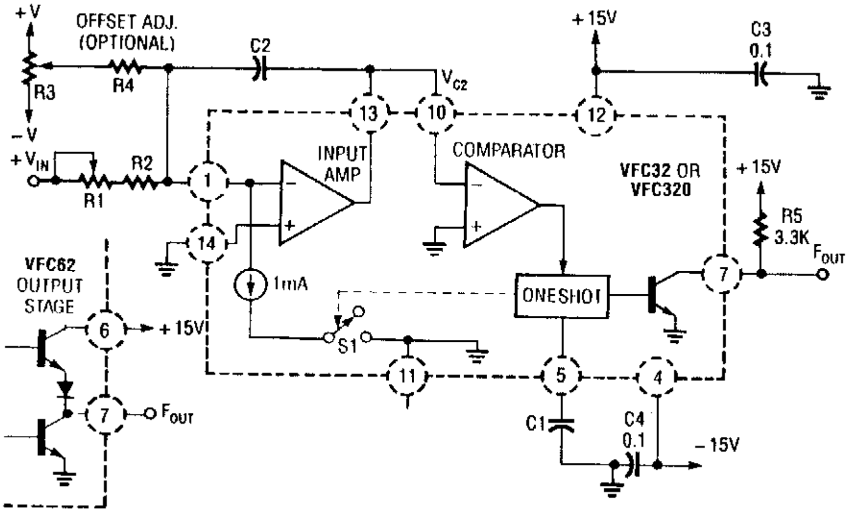
Using a 74HC14 operating at 350 kHz, this voltage tripler delivers approximately 12 V from a 5-V supply.

100

Voltage-To-Frequency Converters

The sources of the following circuits are contained in the Sources section, which begins on page 681. The figure number in the box of each circuit correlates to the entry in the Sources section.

VOLTAGE-TO-FREQUENCY CONVERTER I



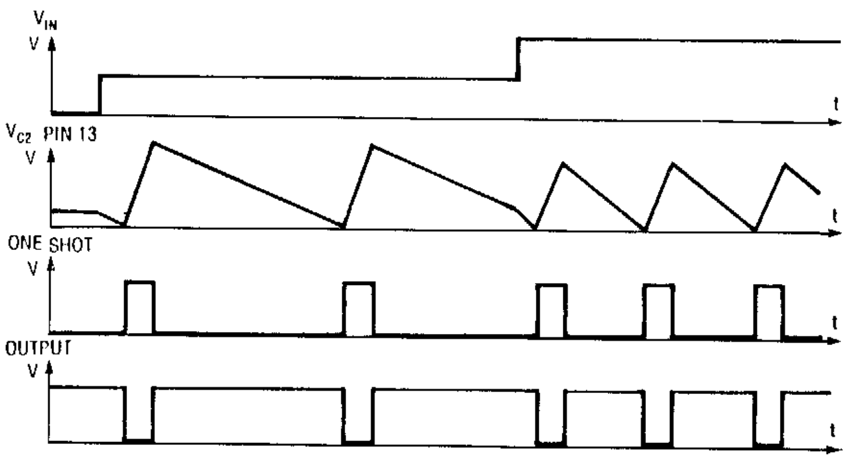
CIRCUIT VALUES

Full-Scale Input	Full-Scale Frequency	C1	C2	R1 (ohms)	R2 (ohms)
1 V	10 kHz	3300 pF	0.01 µF	1K	3K
10 V	10 kHz	3300 pF	0.01 µF	10K	30K
1 V	100 kHz	300 pF	0.001 µF	1K	3K
10 V	100 kHz	300 pF	0.001 µF	10K	30K

RADIO-ELECTRONICS

Fig. 100-1

(a)

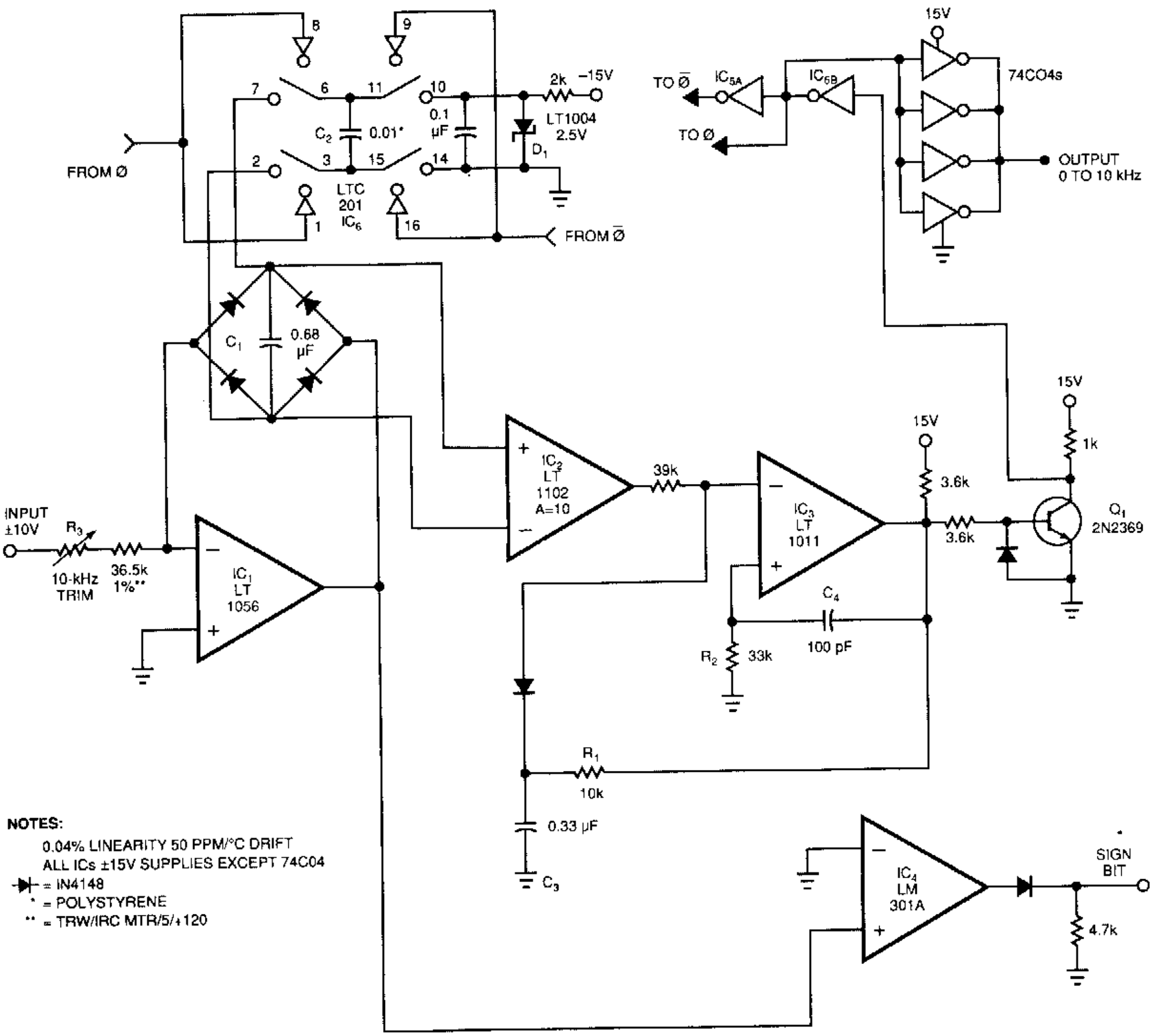


(b)

Using a Burr-Brown VFC 32 IC, this voltage-to-frequency converter uses few components. The circuit values are shown in the figure.

This charge-balanced V/F converter uses a VFC32 or a VFC320 IC. The positive charge from the 1-mA balances the negative charge from the input. V/F converter waveforms are shown in Fig. 100-1(b).

VOLTAGE-TO-FREQUENCY CONVERTER II



EDN

Fig. 100-2

This voltage-to-frequency converter (VFC) accepts the bipolar-ac inputs. For −10- to +10-V inputs, the converter produces a proportional 0- to 10-kHz output. Linearity is 0.04%, and temperature coefficient (TC) measures about 50 ppm/°C.

To understand the circuit, assume that its input sees a bipolar square wave. During the input's positive phase, IC1's output swings negative and drives current through C1 via the full-wave diode bridge. IC1's current causes C1's voltage to ramp up linearity. Instrumentation amplifier IC2 operates at a gain of 10 and measures the differential voltage across C1.

VOLTAGE-TO-FREQUENCY CONVERTER II (*Cont.*)

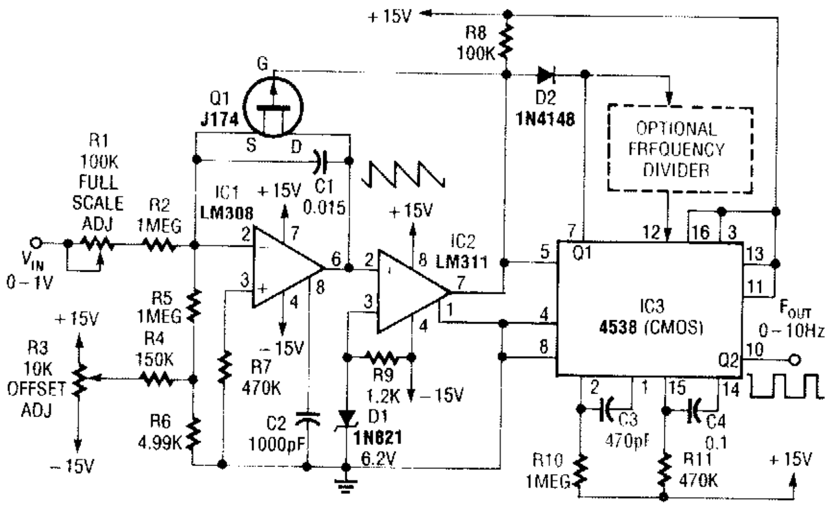
IC2's output biases comparator IC3's negative input. When IC2's output crosses zero, IC3 fires ac positive feedback to IC3's positive input and hangs up IC3's output for about 20 μs. The Q1 level shifter drives ground-referred inverters IC5A and IC5B to deliver biphase drive to LT1004 switch IC6.

IC6, configured as a charge pump, places C2 across C1 each time the inverters switch, which resets C1 to a lower voltage. The LT1004 reference (D1), along with C2's value, determines how much charge the charge pump removes from C1 each time the charge pump cycles. Thus, each time IC2's output tries to cross zero, the charge pump switches C2 across C1, which resets C1 to a small negative voltage and forces IC1 to begin recharging C1.

The frequency of this oscillatory behavior is directly proportional to the input-derived current into IC1. During the time that C1 is ramping toward zero, IC6, places C2 across the reference diode (D1), and prepares C2 for the next discharge cycle.

The action is the same for negative-input excursions, except that IC1's output phasing is reversed. IC2, looking differentially across IC1's diode bridge, sees the same signal as it does for positive inputs; therefore, the circuit's action is identical. IC4, detecting IC1's output polarity, provides a signal bid output.

SIMPLE LOW-FREQUENCY V/F CONVERTER



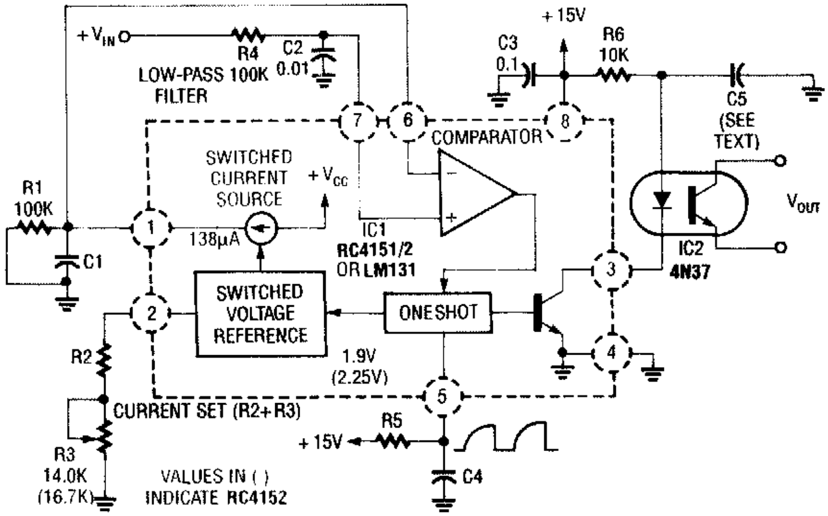
RADIO-ELECTRONICS

Fig. 100-3

In this circuit, C1 is charged to a fixed reference level, then discharged. Integrator IC1 circuit charges C1 until IC1 has −6.2-V output, when comparator IC2 outputs a low. FET Q1, triggers one-section monostable multivibrator IC3, pulls pin 3 low for 470 μs, ensuring that Q1 completely discharges C1. The other section of IC3 produces a longer pulse of about 47 ms.

Full scale of this circuit is 10 Hz. For lower output pulse rates, a counter circuit can be inserted between the sections of IC3. Notice that because C1 does not integrate while Q1 is biased on, this circuit has an error in the output period, which must be as short as possible. Therefore, the circuit's use is limited to low frequencies.

VOLTAGE-TO-FREQUENCY CONVERTER WITH OPTOCOUPLER



CIRCUIT VALUES

Full-Scale Input	Full-Scale Frequency	C1	C4	R5 (ohms)
10 V	1 kHz	10 μF	0.1 μF	6.8K
10 V	10 kHz	1 μF	0.01 μF	6.8K
10 V	100 kHz	0.1 μF	0.001 μF	6.8K

RADIO-ELECTRONICS

Fig. 100-4

In this circuit, a Raytheon RC4151 or National LM131 is used in conjunction with an optocoupler for applications where input-to-output isolation is desirable. Circuit values are shown in the figure for various applications.

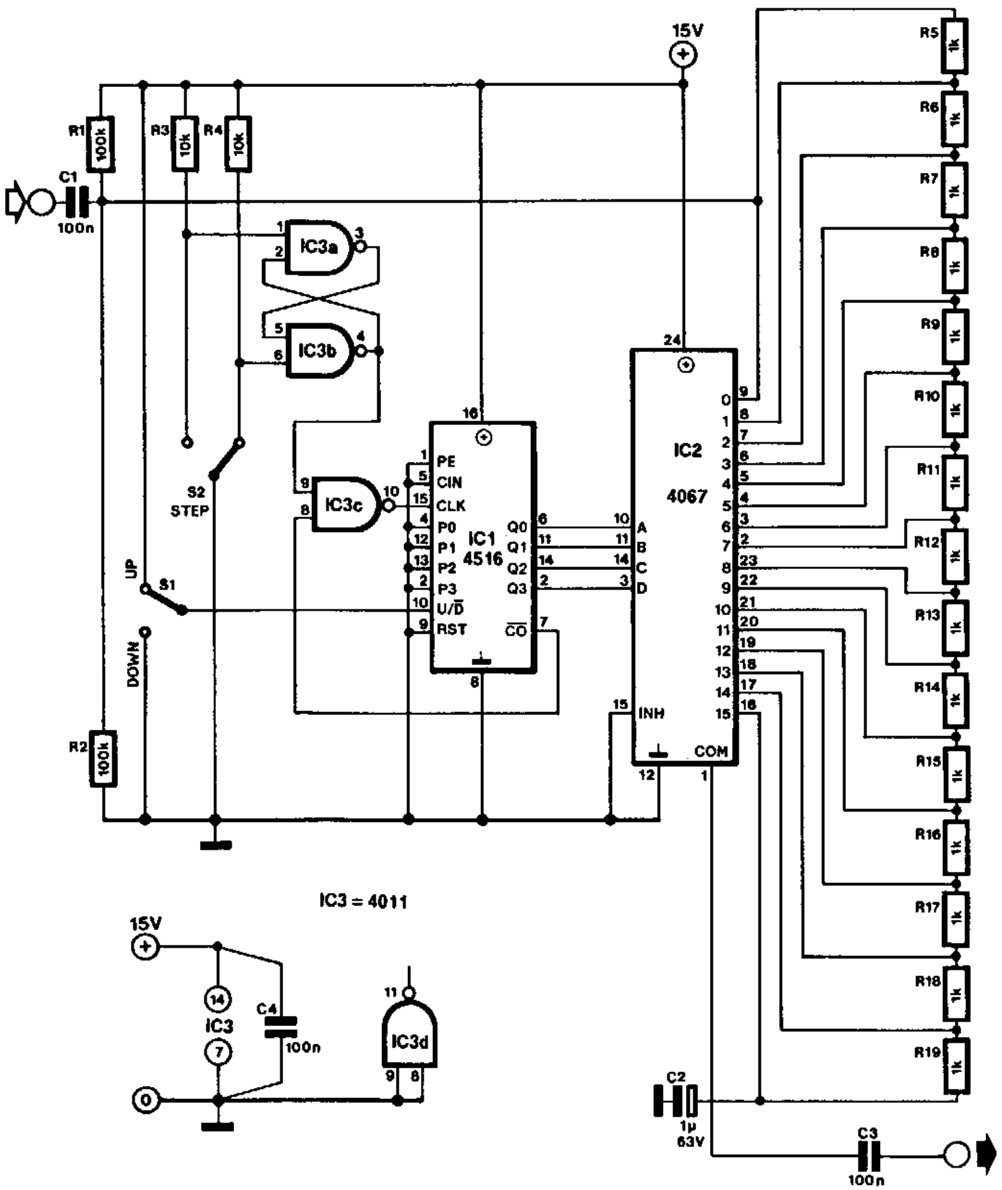
101

Volume/Level-Control Circuit

The source of the following circuit is contained in the Sources section, which begins on page 681. The figure number in the box correlates to the entry in the Sources section.

Digital Volume Control

DIGITAL VOLUME CONTROL



ELEKTOR ELECTRONICS

Fig. 101-1

DIGITAL VOLUME CONTROL (*Cont.*)

The heart of the digitally operated volume control is IC2, a 4067 16-channel analog multiplexer. Depending on the logic state on pins A, B, C, and D of the multiplexer, one of its 16 inputs or outputs is connected to pin 1, which is the "wiper" of the control.

Because a 1-kΩ resistor has been connected between each input and output, the multiplexer can be considered a linear potentiometer with 16 fixed steps. Its overall resistance is 15 kΩ. It is, of course, possible to use a different value for each of the resistors to obtain a different characteristic.

The setting of the potentiometer is controlled by counter IC1. Dependent on the position of switch S1, the counter moves one step up or down when switch S1 is changed over. Circuits IC3A and IC3B debounce S2.

A jump from 0000 to 1111 or the other way around is not possible, because further count pulses are suppressed with the aid of the $\overline{CO}$ line. This line is logic low when both the counter state and signal $U/\overline{D}$ are 0.

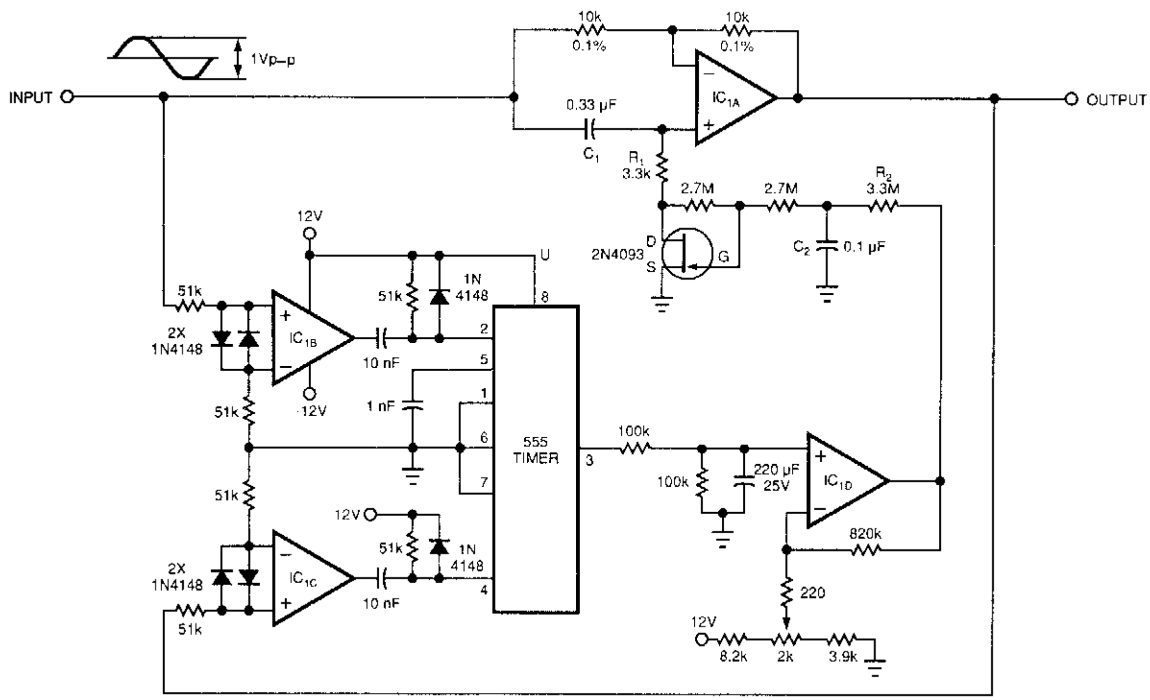
When $U/\overline{D}$ is high and the counter state is 15, $\overline{CO}$ again becomes logic low. It is then necessary to reverse the logic state at $U/\overline{D}$, and thus the direction of counting. The volume control draws a current of around 1 mA.

102

Wave-Shaping Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 681. The figure number in the box of each circuit correlates to the entry in the Sources section.

PHASE SHIFTER



EDN **Fig. 102-1**

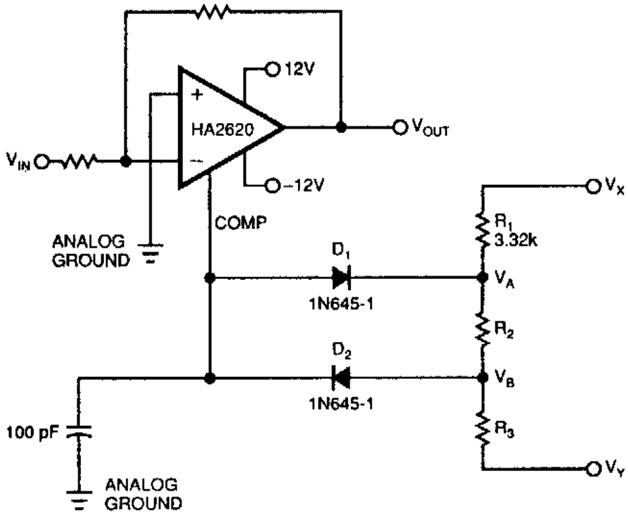
This circuit adds 120 degrees of phase shift to a 50- or 60-Hz input, regardless of the frequency and amplitude fluctuations of that input. The circuit configures a 2N4093 JFET as a voltage-controlled resistor whose value is proportional to the phase difference between the input and the output. The values of C_1, R_1, and r_{DS} determine the amount of phase shift (120° this case.)

A 555 timer implements a phase detector whose two inputs are related to the input and output. The input and output, respectively, drive IC1B and IC1C, which operate as zero-crossing detectors. D1 and D2 limit the positive-going pulses at the 555 inputs. Thus, the falling edges of IC1B and IC1C's outputs control the 555 timer. The timer's output signal stays low for a time that is proportional to the phase shift between the circuit's input and output.

The average value of the timer's output and an offsetting voltage drive IC1D. R2 and C2 filter IC1D's output. The resultant signal controls the JFET. The potentiometer sets the control at a value for which the phase shift between input and output is equal to 120 degrees when the input signal frequency is 50 or 60 Hz. Any differences between the input and output changes the 555 output's average value, thus ultimately modifying the control voltage and the JFET's resistance.

To calibrate the circuit, apply a 50-Hz sine wave with an amplitude of less than 1 Vpp to the input and adjust the potentiometer until the phase shift reads 120° on a digital phase meter. For input frequency variations between 40 and 60 Hz, the phase shift changed by a maximum of ±0.17% (equivalent to an offset of only 0.02°/Hz). The average value at IC1D's noninverting input is 3.864 V.

GLITCH-FREE CLIPPER



EDN

Fig. 102-2

Adding a simple clamping circuit to a Harris 2620 high-speed op amp produces a glitch-free amplifier/clipper. The op amp pin that controls the device's bandwidth is a high-impedance, isolated input. This pin also tracks the device's output voltage.

Therefore, D1, D2, R1, R2, and R3 will clamp the amplifier's output voltage only when the amplifier's input voltage exceeds your clamping-voltage limits. V_D is the diode drop of D1 or D2. The two clamp voltages, $V_A + V_D$ and $V_B - V_D$, are:

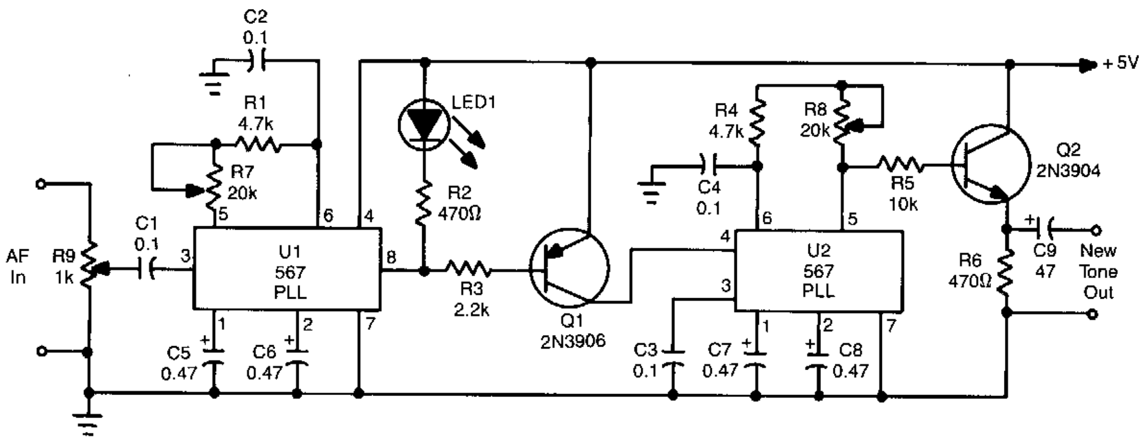
$$V_A = V_X\left(\frac{R_2+R_3}{R_1+R_2+R_3}\right) + V_Y\left(\frac{R_1}{R_1+R_2+R_3}\right)$$

$$V_B = V_X\left(\frac{R_3}{R_1+R_2+R_3}\right) + V_Y\left(\frac{R_1+R_2}{R_1+R_2+R_3}\right)$$

where V_X and V_Y are the clamping circuit's bias voltages. Choosing R_1 lets you determine the values of R_2 and R_3. Try a value for R_1 around 3 kΩ.

One example of this circuit had clamping voltages of ±3.7 V and exhibited THD below −75 dB for a sinusoidal, 30-kHz input signal. When the input signal increased beyond the ±3.7-V clamping voltage, the clipper symmetrically clamped the output voltage with no glitches in the waveform.

SIGNAL CONDITIONER

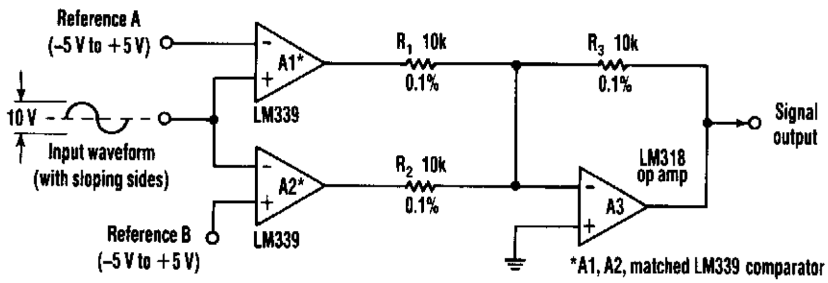


POPULAR ELECTRONICS

Fig. 102-3

This circuit takes audio from a receiver that might have a weak CW or tone signal and uses a PLL (U1) to recover the weak signal. U1 produces a low on receipt of a tone or note of frequency, determined by R1, R7, and C2. The output of U1 (pin 8) goes low, keys tone generator U2, and produces a new tone. The circuit is useful in cleaning up CW reception in static, noise, etc.

HARMONIC GENERATOR



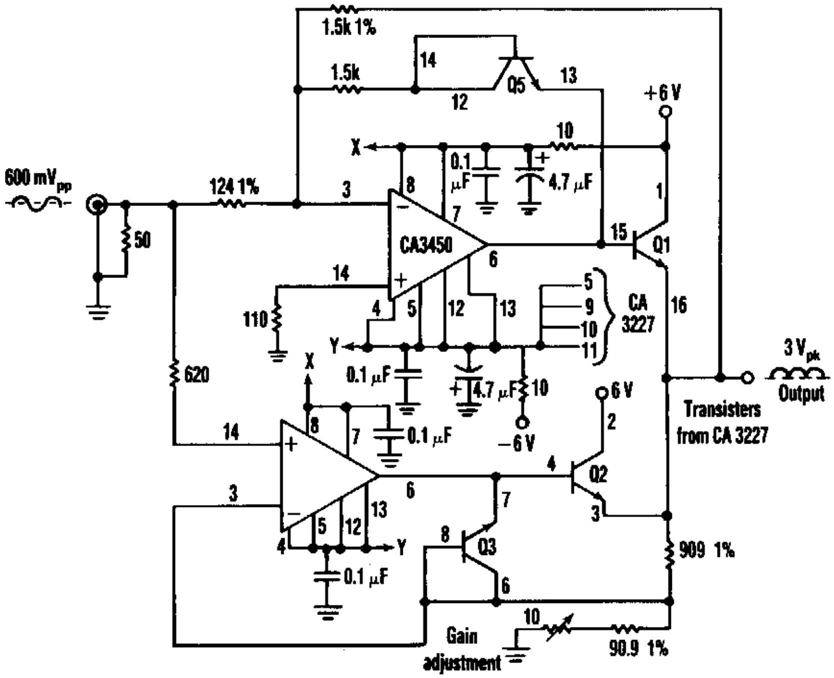
ELECTRONIC DESIGN

Fig. 102-4

This circuit can extract harmonics from various waveforms. With a sloped input waveform, the comparator produces a pulse width that is proportional to a reference plus input amplitude. As the pulse width changes, the harmonics spectrum changes. Combining the two comparator outputs eliminates some harmonics, depending on the duty cycle. Adjusting the references can create virtually any harmonic.

A1 and A2 should be matched, R1 and R2 should be equal to 0.1%, and A3 should have good common-mode rejection and high slew rate.

FULL-WAVE RECTIFIER (TO 10 MHz)

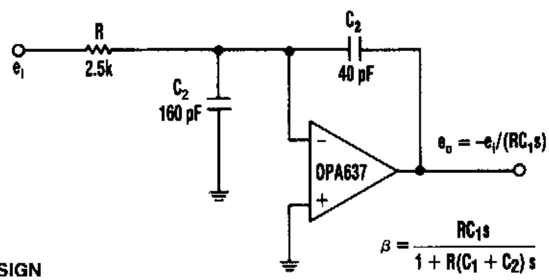


ELECTRONIC DESIGN

Fig. 102-5

Using two CA3450 op amps and a CA3227 transistor array, this circuit will accurately full-wave rectify signals to 10 MHz. Two of the CA3227 transistors drive the output, two are in the feedback circuits. Two transistors serve as clamping diodes, limit the negative-going signals of each amplifier, and keep both amplifiers active during the end cycle. The maximum output is determined by the slew rate of 300 V/μs at highest frequencies. This output equals 300 V/ms $\div 6\pi V$ peak.

CAPACITOR ALLOWS HIGHER SLEW RATES

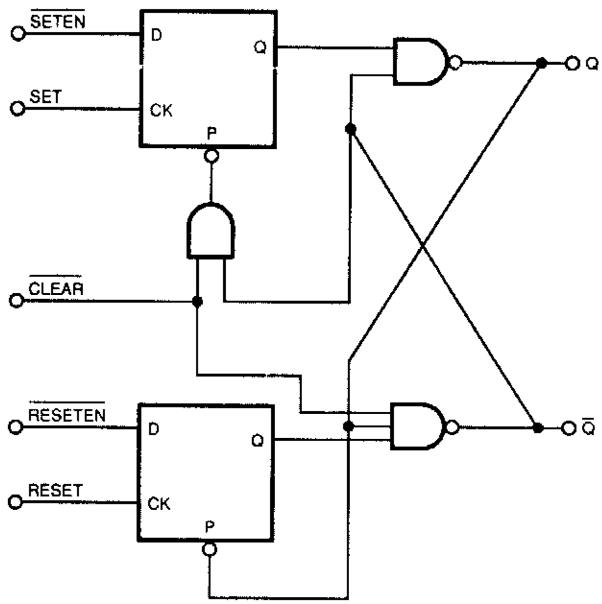


ELECTRONIC DESIGN

Fig. 102-6

In this circuit, a Burr-Brown op amp supplies a slew rate of 135 V/μs. The addition of C2 charges the high-frequency feedback factor to less than unity, and allows higher slew-rate amplifiers to be compensated for greater-than-unity gain.

S/R FLIP-FLOP



EDN

Fig. 102-7

This circuit combines the characteristics of an asynchronous S/R flip-flop and an edge-triggered JK flip-flop. It changes state on the leading edges of its inputs, and ignores the levels at all other times.

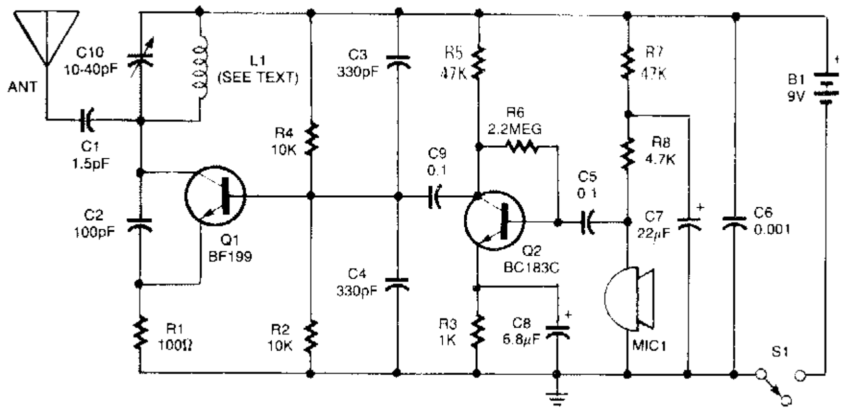
In operation, outputs of both D flip-flops are normally high, going low for brief periods after seeing an edge at their respective clock inputs.

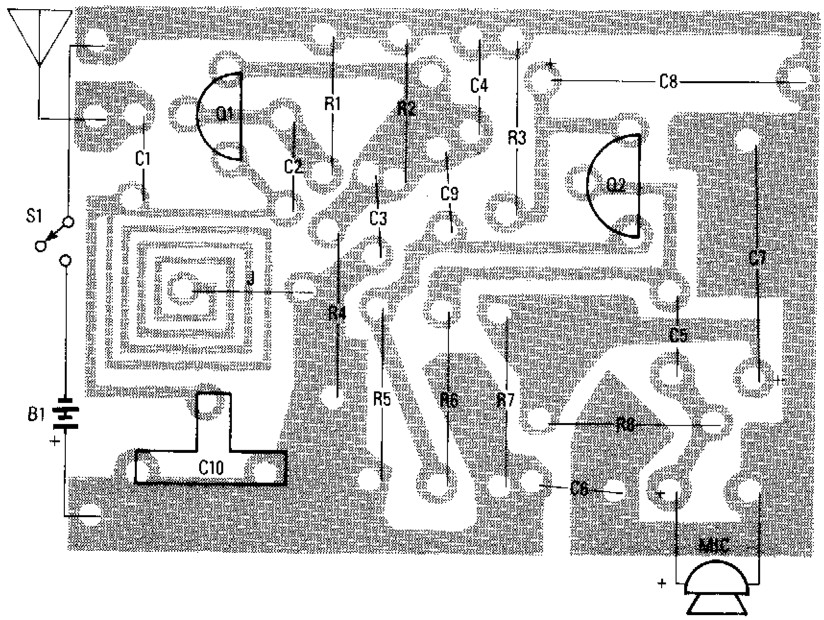
103

Wireless Microphones

The sources of the following circuits are contained in the Sources section, which begins on page 682. The figure number in the box of each circuit correlates to the entry in the Sources section.

WIRELESS FM MICROPHONE



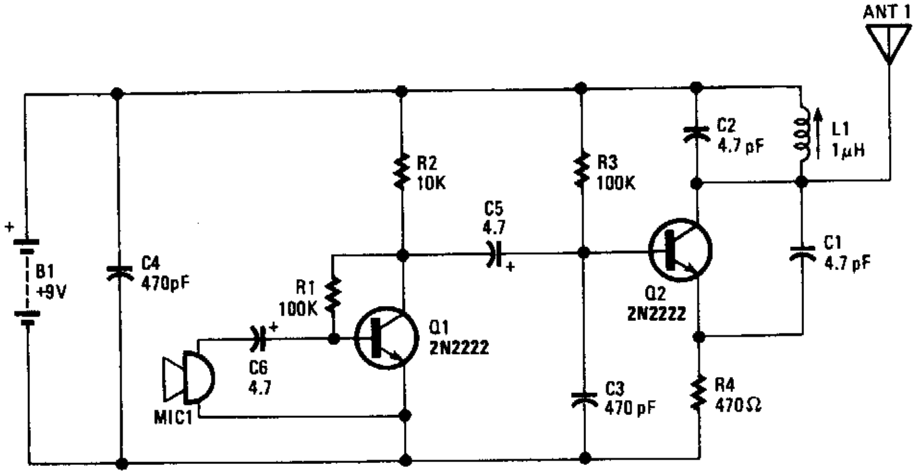


RADIO-ELECTRONICS

Fig. 103-1

A simple FM wireless microphone uses a single BC183C transistor as an audio amplifier. A 2N3565 can be substituted. Q1 is an oscillator that is FM modulated by the signal from Q1. Other transistors can be substituted, but the modulation characteristics should be checked.

WIRELESS MICROPHONE



POPULAR ELECTRONICS

Fig. 103-2

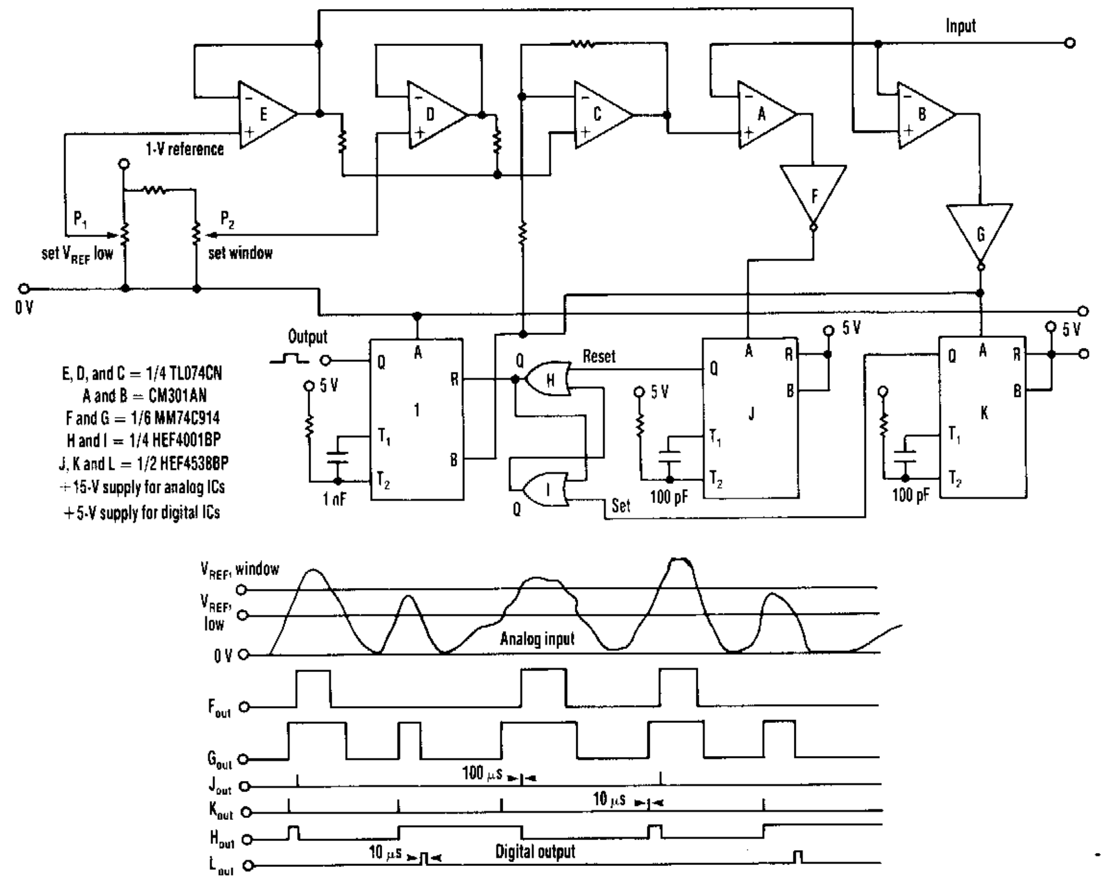
Q1 amplifies the output from an electret microphone MIC1. Audio is fed into oscillator Q2, which modulates the signal. L1C1 is a tank circuit for operation in the 88-MHz region. The antenna is a 6- to 8-inch piece of wire. L1 is a variable inductor in the 1-μH range.

104

Window Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 682. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOW-COST WINDOW COMPARATOR



ELECTRONIC DESIGN

Fig. 104-1

This circuit outputs a TTL-compatible 100-μs pulse whenever the signal falls within the limits set by potentiometers and can be varied to suit the application.

Op amps, E, D, and C are used with the two potentiometers to supply reference voltages, derived from a 1-V precision source, for two other op amps (A and B) configured as voltage comparators. The input signal is taken to the negative inputs of both these comparators. C is wired as a noninverting summing amp, used to derive the higher reference voltage. Consequently, the acceptance window is set. Because the voltage across potentiometer P2 is 0.5 V, the window can be set between 0 and 0.5 V above the value chosen for the lower reference value. The lower value is set by P1.

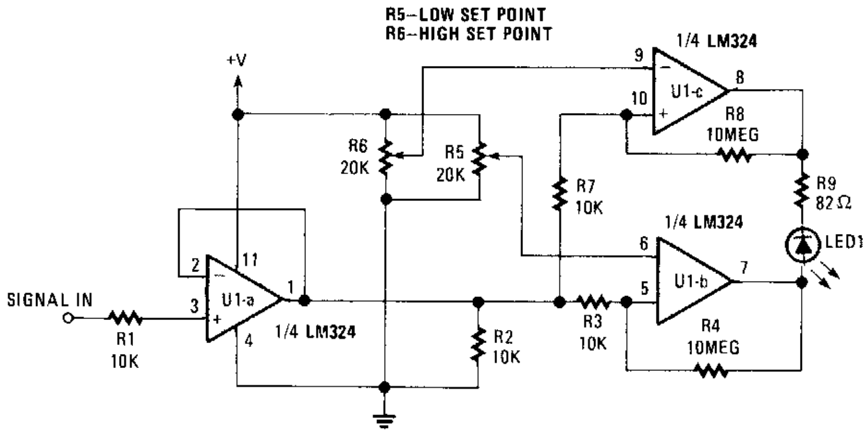
The outputs from the comparators are sent to the inputs of inverting Schmitt triggers F and G. Although these triggers operate from a 5-V supply, they have an extended input-voltage range and are capable of handling the comparator's output voltage swings.

The two monostables, J and K, are triggered on the rising edge of the Schmitt outputs. J and K set and reset the bistable latch formed by two NOR gates. Latch output Q controls the reset of output monostable L, which can only be triggered if its B input goes low, taken from the output of G, while its reset is high.

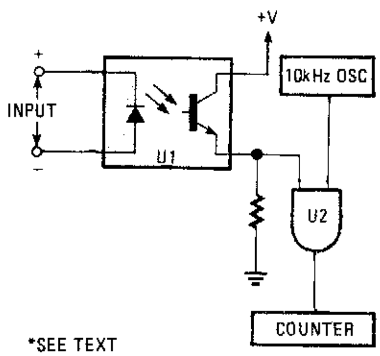
From the timing diagram, when the signal input exceeds the lower reference level, the latch is set and Q goes high. When the higher reference level is exceeded, the latch is reset and Q goes low. If only the lower reference is exceeded, the latch will be set and L is triggered when G's output goes low. When both levels are exceeded, the latch is set, then reset. When G's output goes low, it won't trigger the monostable because its reset is low.

The circuit, as shown, was used for a bandwidth-limited, 500- to 5 000 Hz input signal with an amplitude range of ±1 V. Modifications can easily be made to cover other frequency and amplitude ranges.

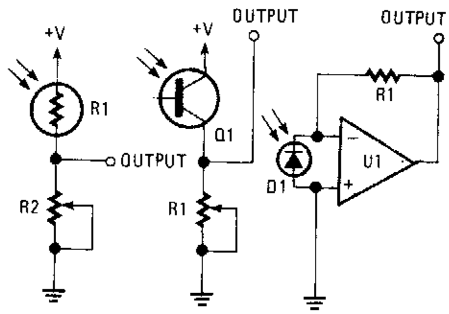
WINDOW GENERATOR



POPULAR ELECTRONICS *Fig. 104-2(a)*



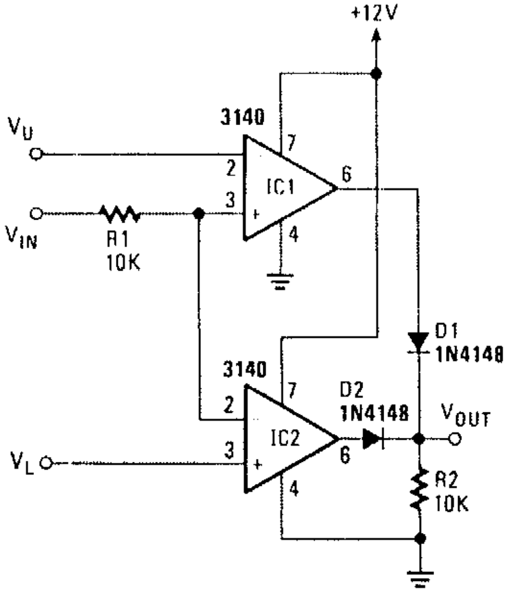
POPULAR ELECTRONICS *Fig. 104-2(b)*



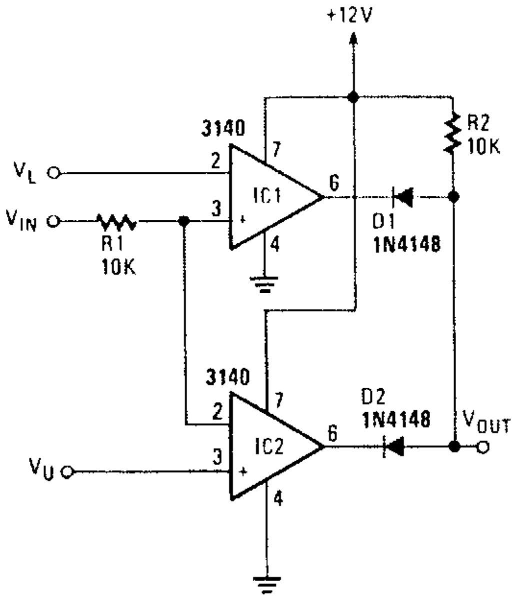
POPULAR ELECTRONICS *Fig. 104-2(c)*

This window generator uses a single LM324 op amp and features two adjustable set points. When the two comparators formed by U1B go high, LED1 lights. When U1C goes high, the LED extinguishes. Some hysteresis is provided by the 10-mΩ resistors. As shown in Fig. 104-2(b), LED1 can be replaced by an optoisolator (etc.) for use in several applications.

WINDOW COMPARATORS



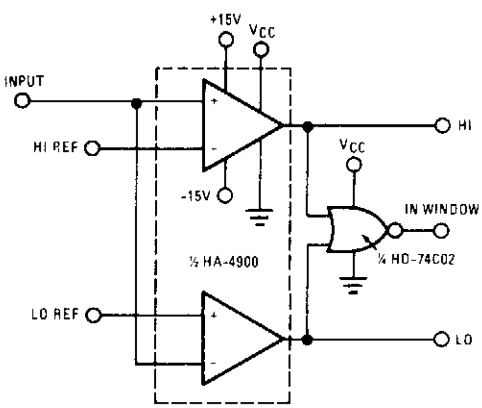
RADIO-ELECTRONICS *Fig. 104-3(a)*



RADIO-ELECTRONICS *Fig. 104-3(b)*

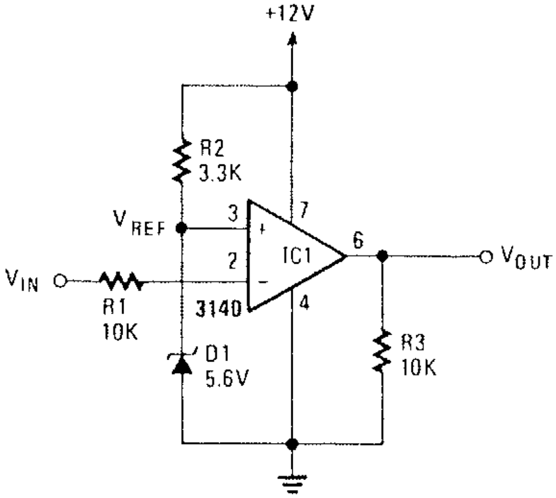
In Fig. 104-3(a), when V_{IN} is between reference voltages V_L and V_U, output V_{OUT} goes low. If $V_{IN} > V_L$, IC2 produces a low. Because IC1 outputs low, if $V_{IN} > V_U$, both outputs are low and V_{OUT} is low. If $V_{IN} < V_L$, both IC1 and IC2 are low. Figure 104-3(b) operates the reverse of this; it produces a high when $V_L < V_{IN} < V_U$.

WINDOW DETECTOR

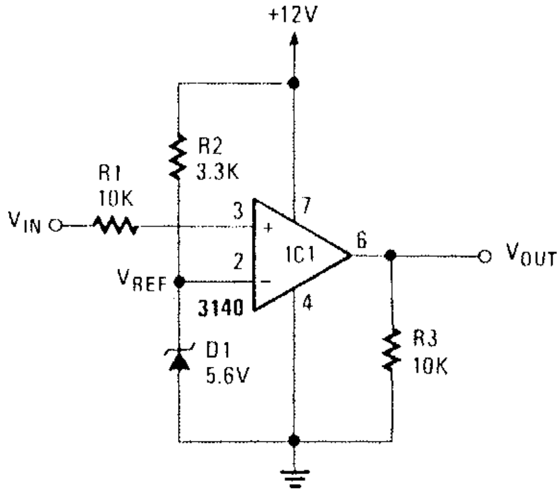


HARRIS *Fig. 104-4*

The high switching speed, low offset current and low offset voltage of the HA-4900 series make this window detector extremely well-suited for applications that require fast, accurate decision-making. This circuit is ideal for industrial-process system-feedback controllers, or "out-of-limit" alarm indicators.



RADIO-ELECTRONICS *Fig. 104-5(a)*



RADIO-ELECTRONICS *Fig. 104-5(b)*

These two comparators are over- and under-voltage comparators. In Fig. 104-5(a), if V_{IN} exceeds the reference voltage, the output of IC1 goes low. In Fig. 104-5(b), if the V_{IN} exceeds the reference, V_{OUT} goes high.

Sources

Chapter 1

Fig. 1-1. Reprinted with permission from Popular Electronics, 11/90, p. 85. © Copyright Gernsback Publications, Inc., 1990.

Fig. 1-2. Reprinted with permission from Popular Electronics, 11/90, © Copyright Gernsback Publications, Inc., 1990.

Fig. 1-3. Elektor Electronics USA, 5/91, p. 51–53.

Fig. 1-4. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 88. © Copyright Gernsback Publications, Inc., 1990.

Chapter 2

Fig. 2-1. Reprinted from EDN, 2/91, p. 184. © 1991 Cahners Publishing Co., a division of Reed Publishing USA.

Chapter 3

Fig. 3-1. Hands-On Electronics, 8/87, p. 94.

Fig. 3-2. Reprinted with permission from Popular Electronics Fact Card No. 185. © Copyright Gernsback Publications, Inc.

Fig. 3-3. Reprinted with permission from Radio-Electronics, 9/87, p. 65. © Copyright Gernsback Publications, Inc., 1987.

Fig. 3-4. Reprinted with permission from Popular Electronics, 12/77, p. 82–83. © Copyright Gernsback Publications, Inc., 1977.

Fig. 3-5. Reprinted with permission from Popular Electronics, 1/91, p. 81. © Copyright Gernsback Publications, Inc., 1991.

Chapter 4

Fig. 4-1. Elektor Electronics USA, 6/91, p. 42.

Fig. 4-2. Reprinted with permission from Popular Electronics, 7/90, p. 80–81. © Copyright Gernsback Publications, Inc., 1990.

Chapter 5

Fig. 5-1. Elektor Electronics, 12/90, p. 523–524.

Fig. 5-2. Hands-On Electronics, 8/87, p. 93.

Fig. 5-3. QST, 5/90, p. 22–24.

Fig. 5-4. Elektor Electronics, 7/89 Supplement, p. 5–6.

Fig. 5-5. Electronics Today International, 4/79, p. 70.

Fig. 5-6. Elektor Electronics, 301 Circuits, p. 30–31.

Fig. 5-7. Electronics Today International, 4/78, p. 31.

Fig. 5-8. Reprinted with permission from Popular Electronics, 4/90, p.37–38. © Copyright Gernsback Publications, Inc., 1990.

Fig. 5-9. Electronics World, 4/66.
Fig. 5-10. Reprinted with permission from Electronic Design, 12/89, p. 71–72. Copyright 1989 Penton Publishing.

Chapter 6

Fig. 6-1. Reprinted with permission from Popular Electronics, 7/91, p. 70–71. © Copyright Gernsback Publications, Inc., 1991.
Fig. 6-2. Hands-On Electronics, 8/87, p. 90.

Chapter 7

Fig. 7-1. Reprinted with permission from Popular Electronics, 7/91, p. 40–42. © Copyright Gernsback Publications, Inc., 1991.
Fig. 7-2. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 90. © Copyright Gernsback Publications, Inc., 1990.
Fig. 7-3. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 23.
Fig. 7-4. Reprinted with permission from Popular Electronics, 3/90. © Copyright Gernsback Publications, Inc., 1990.
Fig. 7-5. Reprinted with permission from Radio-Electronics, 1/90, p. 40. © Copyright Gernsback Publications, Inc., 1990.
Fig. 7-6. Reprinted with permission from Popular Electronics, 10/90, p. 90. © Copyright Gernsback Publications, Inc., 1990.
Fig. 7-7. Reprinted with permission from Radio-Electronics, 9/89, p. 41. © Copyright Gernsback Publications, Inc., 1989.
Fig. 7-8. Reprinted with permission from Radio-Electronics, 1/89, p. 71. © Copyright Gernsback Publications, Inc., 1989.

Chapter 8

Fig. 8-1. Elektor Electronics, 12/90 Supplement, p. 36–37.
Fig. 8-2. Reprinted with permission from Radio-Electronics, 2/90, p. 56. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-3. Reprinted with permission from Radio-Electronics, 2/90, p. 56. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-4. Reprinted with permission from Radio-Electronics, 2/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-5. Reprinted with permission from Radio-Electronics, 2/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-6. Reprinted with permission from Popular Electronics, 8/90, p. 82. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-7. Reprinted with permission from Radio-Electronics, 2/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-8. Reprinted with permission from Radio-Electronics, 2/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-9. Reprinted with permission from Popular Electronics, 3/90, p. 72. © Copyright Gernsback Publications, Inc., 1990.
Fig. 8-10. Elektor Electronics, 12/90, p. 514.
Fig. 8-11. Elektor Electronics, 12/90, p. 511.
Fig. 8-12. Reprinted with permission from Electronic Design, 7/91, p. 139. Copyright 1991, Penton Publishing.
Fig. 8-13. Reprinted with permission from Electronic Design, 9/90, p. 90. Copyright 1990, Penton Publishing.
Fig. 8-14. Reprinted with permission from Popular Electronics, Fact Card No. 173. © Copyright Gernsback Publications, Inc.
Fig. 8-15. QST, 10/90, p. 41.
Fig. 8-16. Reprinted with permission from Popular Electronics, Fact Card No. 185. © Copyright Gernsback Publications, Inc.

Chapter 9

Fig. 9-1. Reprinted with permission from Radio-Electronics, 7/90, p. 33. © Copyright Gernsback Publications, Inc., 1990.
Fig. 9-2. Reprinted with permission from Radio-Electronics, 7/90, p. 36. © Copyright Gernsback Publications, Inc., 1990.
Fig. 9-3. Elektor Electronics, 12/90 Supplement, p. 43.
Fig. 9-4. Reprinted with permission from Radio-Electronics, 7/90, p. 36. © Copyright Gernsback Publications, Inc., 1990.
Fig. 9-5. Reprinted with permission from Radio-Electronics, 7/90, p. 35–36. © Copyright Gernsback Publications, Inc., 1990.
Fig. 9-6. Reprinted with permission from Popular Electronics, 6/89, p. 23. © Copyright Gernsback Publications, Inc., 1989.
Fig. 9-7. Reprinted with permission from Radio-Electronics, 8/90, p. 42. © Copyright Gernsback Publications, Inc., 1990.
Fig. 9-8. Reprinted with permission from Radio-Electronics, 7/90, p. 32. © Copyright Gernsback Publications, Inc., 1990.

Fig. 9-9. Reprinted with permission from Radio-Electronics, 7/90, p. 36. © Copyright Gernsback Publications, Inc., 1990.

Chapter 10

Fig. 10-1. Reprinted with permission from Popular Electronics, 2/90, p. 24–26. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-2. Reprinted with permission from Popular Electronics, 2/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-3. Reprinted with permission from Radio-Electronics, 7/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-4. Reprinted with permission from Popular Electronics, 8/89, p. 24. © Copyright Gernsback Publications, Inc., 1989.

Fig. 10-5. Reprinted with permission from Popular Electronics, 12/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-6. Reprinted with permission from Popular Electronics, 1/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-7. Reprinted with permission from Radio-Electronics, 5/90, p. 65. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-8. Reprinted from EDN, 9/84, p. 290. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 10-9. Reprinted with permission from Radio-Electronics, 5/90, p. 64. © Copyright Gernsback Publications, Inc., 1990.

Fig. 10-10. Reprinted with permission from Popular Electronics, 11/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Chapter 11

Fig. 11-1. Elektor Electronics, 7/89 Supplement, p. 15.

Fig. 11-2. Elektor Electronics, 12/90, p. 527–528.

Fig. 11-3. Reprinted with permission from Popular Electronics, 12/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.

Fig. 11-4. Reprinted with permission from Popular Electronics, 9/89, p. 22. © Copyright Gernsback Publications, Inc., 1989.

Fig. 11-5. Reprinted with permission from Hands-On Electronics/Popular Electronics, 12/88, p. 24. © Copyright Gernsback Publications, Inc., 1988.

Fig. 11-6. Elektor Electronics, 7/89, p. S41.

Fig. 11-7. Reprinted with permission from Popular Electronics, 7/89, p. 76–77. © Copyright Gernsback Publications, Inc., 1989.

Fig. 11-8. Reprinted with permission from Popular Electronics, 8/89, p. 23. © Copyright Gernsback Publications, Inc., 1989.

Fig. 11-9. Hands-On Electronics, 9–10/86, p. 26.

Fig. 11-10. Reprinted with permission from Popular Electronics, 12/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.

Chapter 12

Fig. 12-1. Reprinted with permission from Radio-Electronics, 4/90, p. 47. © Copyright Gernsback Publications, Inc., 1990.

Fig. 12-2. Elektor Electronics, 12/90, p. 57.

Fig. 12-3. Reprinted with permission from Radio-Electronics, 3/89, p. 51–52. © Copyright Gernsback Publications, Inc., 1989.

Fig. 12-4. Elektor Electronics, 12/90, p. 526.

Fig. 12-5. Reprinted with permission from Radio-Electronics, 8/91, p. 8. © Copyright Gernsback Publications, Inc., 1991.

Chapter 13

Fig. 13-1. Elektor Electronics, 12/90, p. 522.

Fig. 13-2. Hands-On Electronics, 11/87, p. 92.

Fig. 13-3. Reprinted with permission from Popular Electronics, Fact Card No. 134. © Copyright Gernsback Publications, Inc.

Fig. 13-4. Reprinted with permission from Popular Electronics, Fact Card No. 95. © Copyright Gernsback Publications, Inc.

Fig. 13-5. Reprinted with permission from Popular Electronics, 3/90, p. 38. © Copyright Gernsback Publications, Inc., 1990.

Fig. 13-6. Reprinted from EDN, 2/7/85, p. 243. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 14

Fig. 14-1. Elektor Electronics, 12/90, p. 519.

Fig. 14-2. Reprinted from EDN, 5/91, p. 165. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 14-3. Reprinted from EDN, 6/91, p. 168. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 14-4. Reprinted from EDN, 5/90, p. 148. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 14-5. Reprinted from EDN, 2/91, p. 108. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 14-6. Elektor Electronics, 5/91, p. 33.
Fig. 14-7. Reprinted with permission from Popular Electronics, 8/90, p. 40–41. © Copyright Gernsback Publications, Inc., 1990.
Fig. 14-8. 42 New Ideas, 1984, p. 14, Gernsback Publications, Inc.
Fig. 14-9. Electronic Engineering, 11/78, p. 24.
Fig. 14-10. Maxim, Seminar Applications Book, 1988/89, p. 35.
Fig. 14-11. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 19.

Chapter 15

Fig. 15-1. Reprinted from EDN, Design Ideas Special Issue, Vol. III, 2/89, p. 47. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 15-2. Reprinted from EDN, Design Ideas Special Issue, Vol. IV, 7/20/89, p. 26. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 15-3. 73 Amateur Radio, 4/91, p. 11–12.

Chapter 16

Fig. 16-1. Reprinted with permission from Radio-Electronics, 5/90, p. 64. © Copyright Gernsback Publications, Inc., 1990.
Fig. 16-2. Reprinted with permission from Popular Electronics, 8/90, p. 92. © Copyright Gernsback Publications, Inc., 1990.
Fig. 16-3. Reprinted with permission from Popular Electronics, 12/90, p. 41–42. © Copyright Gernsback Publications, Inc., 1990.
Fig. 16-4. Reprinted with permission from Popular Electronics, Fact Card No. 173. © Copyright Gernsback Publications, Inc.
Fig. 16-5. Reprinted with permission from Radio-Electronics, 5/90, p. 65. © Copyright Gernsback Publications, Inc., 1990.
Fig. 16-6. Reprinted with permission from Radio-Electronics, 5/90, p. 62. © Copyright Gernsback Publications, Inc., 1990.
Fig. 16-7. Reprinted with permission from Radio-Electronics, 5/90, p. 62. © Copyright Gernsback Publications, Inc., 1990.

Chapter 17

Fig. 17-1. Reprinted from EDN, 2/91, p. 105. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 17-2. Reprinted from EDN, 11/90, p. 282. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 17-3. Hands-On Electronics, Fact Card No. 89.

Chapter 18

Fig. 18-1. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 158. © Copyright Gernsback Publications, Inc., 1990.
Fig. 18-2. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 162. © Copyright Gernsback Publications, Inc., 1990.

Chapter 19

Fig. 19-1. Reprinted from EDN, 1/91, p. 160–161. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 19-2. Elektor Electronics, 12/90, p. 534–535.

Chapter 20

Fig. 20-1. Reprinted from EDN, 3/91, p. 161. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-2. Elektor Electronics, 6/89, p. S42.
Fig. 20-3. Reprinted from EDN, 9/89, p. 168. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-4. Reprinted from EDN, 5/91, p. 166. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-5. Electronic Engineering, 2/90, p. 29.
Fig. 20-6. Reprinted from EDN, 2/90, p. 135. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-7. Reprinted with permission from Popular Electronics, 8/91, p. 20. © Copyright Gernsback Publications, Inc., 1991.

Fig. 20-8. Reprinted with permission from Popular Electronics, 1/91, p. 23–24. © Copyright Gernsback Publications, Inc., 1991.
Fig. 20-9. Reprinted with permission from Electronic Design, 9/89, p. 93. Copyright 1989, Penton Publishing.
Fig. 20-10. Reprinted with permission from Popular Electronics, 12/90, p. 23. © Copyright Gernsback Publications, Inc., 1990.
Fig. 20-11. Elektor Electronics, 7/89, p. S36.
Fig. 20-12. Reprinted from EDN, Design Ideas Special Issue, Vol. IV, 7/20/89, p. 29. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-13. Reprinted from EDN, 1/90, p. 140. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-14. Reprinted from EDN, 11/90, p. 22. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 20-15. Harris, Analog Product Data Book, 1988, p.10–23.
Fig. 20-16. Reprinted with permission from Popular Electronics, 1/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Chapter 21

Fig. 21-1. Reprinted from EDN, 7/90, p. 192. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 21-2. Reprinted from EDN, 7/89, p. 151. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 21-3. Elektor Electronics, 12/90, p. 513–514.
Fig. 21-4. Elektor Electronics, 12/90, p. 56.
Fig. 21-5. Reprinted with permission from Electronic Design, 6/91, p. 110. Copyright 1991, Penton Publishing.
Fig. 21-6. Elektor Electronics, 12/90, p. 530–531.
Fig. 21-7. Reprinted from EDN, 3/90, p. 140. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 21-8. Linear Technology, Design Note No. 2.
Fig. 21-9. Reprinted with permission from Electronic Design, 9/90, p. 93.
Fig. 21-10. Linear Technology, 6/27/91, Advertisement.
Fig. 21-11. Linear Technology, 6/27/91, Advertisement.
Fig. 21-12. Maxim, 1986 Power Supply Circuits, p. 26.
Fig. 21-13. Reprinted with permission from Radio-Electronics, 6/86, p. 59. © Copyright Gernsback Publications, Inc., 1986.
Fig. 21-14. Reprinted with permission from Radio-Electronics, 9/89, p. 60. © Copyright Gernsback Publications, Inc., 1989.

Chapter 22

Fig. 22-1. Reprinted from EDN, 10/90, p. 132. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 22-2. RF Design, 7/90, p. 63.
Fig. 22-3. RF Design, 7/90, p. 63.
Fig. 22-4. Reprinted with permission from Popular Electronics, 11/91, p. 68. © Copyright Gernsback Publications, Inc., 1991.
Fig. 22-5. Reprinted with permission from Popular Electronics, 7/91, p. 58. © Copyright Gernsback Publications, Inc., 1991.
Fig. 22-6. 73 Amateur Radio, 10/90, p. 65.
Fig. 22-7. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-8. RF Design, 7/90, p. 64.
Fig. 22-9. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-10. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-11. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-12. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-13. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-14. Reprinted with permission from Popular Electronics, 11/90, p. 69. © Copyright Gernsback Publications, Inc., 1990.
Fig. 22-15. Valpey-Fisher Corp., A User's Guide to Quartz Crystal Oscillators.
Fig. 22-16. Reprinted from EDN, 11/90, p. 234. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 22-17. Elektor Electronics, 302 Circuits, p. 128–129.
Fig. 22-18. William Sheets.

Chapter 23

Fig. 23-1. Reprinted from EDN, 12/89, p. 262. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 23-2. Harris, Analog Product Data Book, 1988, p. 10–49.

Fig. 23-3. Harris, Analog Product Data Book, 1988, p. 2-105.
Fig. 23-4. Reprinted with permission from Radio-Electronics, 12/89, p. 58. © Copyright Gernsback Publications, Inc., 1989.
Fig. 23-5. Reprinted with permission from Electronic Design, 6/90, p. 102-103. Copyright 1991, Penton Publishing.

Chapter 24

Fig. 24-1. Elektor Electronics, 12/90, p. 520.
Fig. 24-2. Reprinted from EDN, 12/89, p. 260-261. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 24-3. Reprinted with permission from Popular Electronics, 8/91, p. 73. © Copyright Gernsback Publications, Inc., 1991.
Fig. 24-4. Reprinted with permission from Popular Electronics, 3/91, p. 92. © Copyright Gernsback Publications, Inc., 1991.
Fig. 24-5. Reprinted from EDN, 5/90, p. 173. Copyright 1991, Penton Publishing.
Fig. 24-6. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.
Fig. 24-7. ARRL Handbook, 1991, p. 18-13.
Fig. 24-8. Reprinted with permission from Popular Electronics, 9/91, p. 75. © Copyright Gernsback Publications, Inc., 1991.
Fig. 24-9. RCA, ICAN #6629, Design Guide for Fire Detection Systems, p. 6.
Fig. 24-10. Reprinted with permission from Popular Electronics, 5/91, p. 63. © Copyright Gernsback Publications, Inc., 1991.
Fig. 24-11. RF Design, 4/90, p. 35-36.
Fig. 24-12. Reprinted with permission from Electronic Design, 12/90, p. 61. Copyright 1991, Penton Publishing.
Fig. 24-13. Reprinted with permission from Popular Electronics, 3/90, p. 46. © Copyright Gernsback Publications, Inc., 1990.
Fig. 24-14. ARRL Handbook, 1991, p. 18-14.
Fig. 24-15. Harris, Analog Product Data Book, 1988, p. 10-84.
Fig. 24-16. Reprinted with permission from Popular Electronics, 9/91, p. 74. © Copyright Gernsback Publications, Inc., 1991.
Fig. 24-17. Reprinted with permission from Popular Electronics, 9/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.
Fig. 24-18. Reprinted with permission from Radio-Electronics, 9/89, p. 59. © Copyright Gernsback Publications, Inc., 1989.

Chapter 25

Fig. 25-1. Reprinted from EDN, 12/90, p. 228-230. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 25-2. 73 Amateur Radio, 7/90, p. 9-11.

Chapter 26

Fig. 26-1. Reprinted with permission from Electronic Design, 6/91, p. 109-114. Copyright 1991, Penton Publishing.
Fig. 26-2. Reprinted with permission from Electronic Design, 3/91, p. 87-88. Copyright 1991, Penton Publishing.
Fig. 26-3. Reprinted from EDN, 6/89, p. 207. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 26-4. 73 Amateur Radio, 3/91, p. 74.
Fig. 26-5. Reprinted with permission from Electronic Design, 11/89, p. 111. Copyright 1989, Penton Publishing.
Fig. 26-6. Reprinted from EDN, 6/91, p. 174. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 27

Fig. 27-1. Elektor Electronics, 12/90, p. 526-527.
Fig. 27-2. Reprinted with permission from Electronic Design, 12/88, p. 104. Copyright 1991, Penton Publishing.
Fig. 27-3. Reprinted with permission from Radio-Electronics, 9/87, p. 65. © Copyright Gernsback Publications, Inc., 1987.

Chapter 28

Fig. 28-1. Reprinted with permission from Popular Electronics, 11/90, p. 61-62. © Copyright Gernsback Publications, Inc., 1990
Fig. 28-2. Reprinted with permission from Popular Electronics, 12/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Chapter 29

Fig. 29-1. 73 Amateur Radio, 2/91, p. 24.
Fig. 29-2. Reprinted with permission from Popular Electronics, 10/89, p. 26. © Copyright Gernsback Publications, Inc., 1989.
Fig. 29-3. Elektor Electronics, 302 Circuits, p. 134.

Chapter 30

Fig. 30-1. Electronic Engineering, 2/90, p. 30.
Fig. 30-2. Reprinted with permission from Popular Electronics, 11/89, p. 101. © Copyright Gernsback Publications, Inc., 1989.
Fig. 30-3. Elektor Electronics, 12/90, p. 530.
Fig. 30-4. Elektor Electronics, 7/89, p. S37.
Fig. 30-5. ARRL Handbook, 1991, p. 16–6.
Fig. 30-6. Reprinted from EDN, 2/91, p. 186–188. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 30-7. Reprinted with permission from Radio-Electronics, 2/90, p. 58. © Copyright Gernsback Publications, Inc., 1990.
Fig. 30-8. Reprinted with permission from Radio-Electronics, 5/89, p. 58. © Copyright Gernsback Publications, Inc., 1989.
Fig. 30-9. Reprinted with permission from Radio-Electronics, 5/89, p. 58. © Copyright Gernsback Publications, Inc., 1989.
Fig. 30-10. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 52.
Fig. 30-11. Reprinted with permission from Radio-Electronics, 5/89, p. 58. © Copyright Gernsback Publications, Inc., 1989.
Fig. 30-12. Reprinted with permission from Radio-Electronics, 2/90, p. 58. © Copyright Gernsback Publications, Inc., 1990.
Fig. 30-13. Reprinted with permission from Radio-Electronics, 2/90, p. 58. © Copyright Gernsback Publications, Inc., 1990.
Fig. 30-14. Electronics Today International, 4/79, p. 76.
Fig. 30-15. Reprinted with permission from Radio-Electronics, 10/90, p. 78. © Copyright Gernsback Publications, Inc., 1990.

Chapter 31

Fig. 31-1. Reprinted from EDN, 10/88, p. 327. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 31-2. Reprinted from EDN, 6/84, p. 242. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 31-3. Reprinted with permission from Radio-Electronics, 2/91, p. 33–34. © Copyright Gernsback Publications, Inc., 1991.
Fig. 31-4. Reprinted with permission from Radio-Electronics, 2/87, p. 36. © Copyright Gernsback Publications, Inc., 1987.
Fig. 31-5. Reprinted from EDN, 7/73, p. 87. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 31-6. Reprinted with permission from Popular Electronics, 1/91, p. 25. © Copyright Gernsback Publications, Inc., 1991.
Fig. 31-7. Reprinted from EDN, 7/73, p. 87. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 31-8. Elektor Electronics, 12/90 Supplement, p. 43.
Fig. 31-9. Reprinted with permission from Popular Electronics, 6/89, p. 22. © Copyright Gernsback Publications, Inc., 1989.

Chapter 32

Fig. 32-1. Reprinted with permission from Popular Electronics, 3/91, p. 24. © Copyright Gernsback Publications, Inc., 1991.
Fig. 32-2. Reprinted from EDN, 12/90, p. 217. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 32-3. Elektor Electronics, 301 Circuits, p. 131.
Fig. 32-4. Elektor Electronics, 12/90, p. 55–56.
Fig. 32-5. Reprinted with permission from Popular Electronics, 9/91, p. 76. © Copyright Gernsback Publications, Inc., 1991.
Fig. 32-6. Reprinted with permission from Radio-Electronics, 7/90, p. 59. © Copyright Gernsback Publications, Inc., 1990.
Fig. 32-7. Reprinted with permission from Popular Electronics, Fact Card No. 107. © Copyright Gernsback Publications, Inc.
Fig. 32-8. Reprinted with permission from Popular Electronics, 3/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.
Fig. 32-9. Reprinted with permission from Popular Electronics, 11/90, p. 22. © Copyright Gernsback Publications, Inc., 1990.
Fig. 32-10. Reprinted with permission from Popular Electronics, 1/90, p. 43. © Copyright Gernsback Publications, Inc., 1990.
Fig. 32-11. Reprinted with permission from Popular Electronics, 1/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Chapter 33

Fig. 33-1. Reprinted with permission from Radio-Electronics, 6/91, p. 56. © Copyright Gernsback Publications, Inc., 1991.

Fig. 33-2. Reprinted with permission from Popular Electronics, Fact Card No. 134. © Copyright Gernsback Publications, Inc.
Fig. 33-3. Elektor Electronics, 12/90, p. 510 – 511.
Fig. 33-4. Reprinted with permission from Popular Electronics, Fact Card No. 98. © Copyright Gernsback Publications, Inc.
Fig. 33-5. Reprinted with permission from Radio-Electronics, 6/91, p. 58 – 59. © Copyright Gernsback Publications, Inc., 1991.

Chapter 34

Fig. 34-1. Hands-On Electronics, 11/86, p. 33.
Fig. 34-2. Reprinted with permission from Electronic Design, 5/91, p. 95. Copyright 1991, Penton Publishing.
Fig. 34-3. Reprinted with permission from Radio-Electronics, 2/91, p. 23. © Copyright Gernsback Publications, Inc., 1991.
Fig. 34-4. Reprinted with permission from Radio-Electronics, 6/91, p. 56 – 57. © Copyright Gernsback Publications, Inc., 1991.
Fig. 34-5. Reprinted with permission from Radio-Electronics, 12/89, p. 57. © Copyright Gernsback Publications, Inc., 1989.
Fig. 34-6. Reprinted with permission from Electronic Design, 8/90, p. 94 – 96. Copyright 1991, Penton Publishing.
Fig. 34-7. Hands-On Electronics, 8/87, p. 93.
Fig. 34-8. Reprinted with permission from Electronic Design, 6/90, p. 100. Copyright 1991, Penton Publishing.
Fig. 34-9. Reprinted with permission from Electronic Design, 9/90, p. 89. Copyright 1991, Penton Publishing.

Chapter 35

Fig. 35-1. Reprinted with permission from Radio-Electronics, 4/90, p. 33 – 36. © Copyright Gernsback Publications, Inc., 1990.
Fig. 35-2. Reprinted with permission from Popular Electronics, 6/89, p. 90. © Copyright Gernsback Publications, Inc., 1989.
Fig. 35-3. Reprinted with permission from Radio-Electronics, 12/89, p. 57. © Copyright Gernsback Publications, Inc., 1989.
Fig. 35-4. Reprinted with permission from Popular Electronics, 10/77, p. 82. © Copyright Gernsback Publications, Inc., 1977.
Fig. 35-5. Reprinted with permission from Popular Electronics, 1/90, p. 70. © Copyright Gernsback Publications, Inc., 1990.
Fig. 35-6. Reprinted with permission from Popular Electronics, 12/89, p. 24 – 25. © Copyright Gernsback Publications, Inc., 1989.

Chapter 36

Fig. 36-1. Reprinted from EDN, 10/90, p. 235. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 37

Fig. 37-1. Electronic Engineering, 4/90, p. 42.
Fig. 37-2. Elektor Electronics, 12/90, p. 532 – 533.
Fig. 37-3. Reprinted with permission from Popular Electronics, 9/91, p. 22. © Copyright Gernsback Publications, Inc., 1991.
Fig. 37-4. 73 Amateur Radio, 10/90, p. 46.
Fig. 37-5. Reprinted with permission from Popular Electronics, 3/91, p. 25. © Copyright Gernsback Publications, Inc., 1991.
Fig. 37-6. Reprinted with permission from Radio-Electronics, 2/91, p. 83. © Copyright Gernsback Publications, Inc., 1991.
Fig. 37-7. Elektor Electronics, 7/89, p. S44.
Fig. 37-8. Elektor Electronics, 7/89, p. 40 – 41.
Fig. 37-9. Elektor Electronics, 7/89, p. S41.
Fig. 37-10. Reprinted with permission from Radio-Electronics, 7/90, p. 59. © Copyright Gernsback Publications, Inc., 1990.
Fig. 37-11. Ham Radio, 4/90, p. 66.

Chapter 38

Fig. 38-1. Reprinted with permission from R-E Experimenter's Handbook, 1987, p. 59. © Copyright Gernsback Publications, Inc., 1987.
Fig. 38-2. Reprinted with permission from Popular Electronics, 4/90, p. 91. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-3. Reprinted with permission from Radio-Electronics, 4/89, p. 47. © Copyright Gernsback Publications, Inc., 1989.
Fig. 38-4. Reprinted with permission from Popular Electronics, 9/90, p. 23. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-5. Reprinted with permission from R-E Experimenter's Handbook, 1987, p. 59. © Copyright Gernsback Publications, Inc., 1987.

Fig. 38-6. Reprinted with permission from Popular Electronics, 1/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-7. Reprinted with permission from Radio-Electronics, 4/91, p. 73. © Copyright Gernsback Publications, Inc., 1991.
Fig. 38-8. Reprinted with permission from Popular Electronics, 2/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-9. Reprinted with permission from Popular Electronics/Hands-On Electronics, 4/89, p. 83. © Copyright Gernsback Publications, Inc., 1989.
Fig. 38-10. Reprinted with permission from Popular Electronics, 4/90, p. 90 – 91. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-11. Reprinted with permission from Popular Electronics, 3/90, p. 28. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-12. Reprinted with permission from Popular Electronics, 12/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.
Fig. 38-13. Reprinted with permission from Popular Electronics, 2/82, p. 99. © Copyright Gernsback Publications, Inc., 1982.

Chapter 39

Fig. 39-1. Elektor Electronics, 12/90, p. 529.
Fig. 39-2. Reprinted from EDN, 2/91, p. 105. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 39-3. DATEL, Data Conversion Components, p. 4 – 17.
Fig. 39-4. Reprinted with permission from Electronic Design, 12/88, p. 67. Copyright 1988, Penton Publishing.

Chapter 40

Fig. 40-1. Elektor Electronics, 12/90, p. 515 – 516.

Chapter 41

Fig. 41-1. Reprinted with permission from Electronic Design, 8/90, p. 94. Copyright 1990, Penton Publishing.
Fig. 41-2. Elektor Electronics, 12/90, p. 512.
Fig. 41-3. Reprinted with permission from Electronic Design, 4/91, p. 102. Copyright 1991, Penton Publishing.
Fig. 41-4. Harris, Analog Product Data Book, 1988, p. 2 – 105.

Chapter 42

Fig. 42-1. 73 Amateur Radio, 5/91, p. 14 – 15.
Fig. 42-2. QST, 6/91, p. 45.
Fig. 42-3. Reprinted by permission of Texas Instruments, Linear and Interface Circuits Applications, 1985, Vol. 1, p. 4 – 3.

Chapter 43

Fig. 43-1. Reprinted with permission from Popular Electronics, 8/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.
Fig. 43-2. Reprinted with permission from Popular Electronics/Hands-On Electronics, 1/89, p. 44. © Copyright Gernsback Publications, Inc., 1989.
Fig. 43-3. Elektor Electronics, 7/89, p. 38.
Fig. 43-4. Reprinted with permission from Popular Electronics, 8/89, p. 23. © Copyright Gernsback Publications, Inc., 1989.
Fig. 43-5. Elektor Electronics, 301 Circuits, p. 8.
Fig. 43-6. Reprinted with permission from Popular Electronics, 8/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.
Fig. 43-7. Reprinted with permission from Popular Electronics, 8/90, p. 23. © Copyright Gernsback Publications, Inc., 1990.
Fig. 43-8. Reprinted with permission from Popular Electronics, 11/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.
Fig. 43-9. Reprinted with permission from Popular Electronics, 9/90, p. 26. © Copyright Gernsback Publications, Inc., 1990.
Fig. 43-10. Reprinted with permission from Popular Electronics, 9/90, p. 26. © Copyright Gernsback Publications, Inc., 1990.
Fig. 43-11. Elektor Electronics, 12/90, p. 58 – 59.
Fig. 43-12. Reprinted with permission from Popular Electronics, 9/91, p. 19. © Copyright Gernsback Publications, Inc., 1991.
Fig. 43-13. Reprinted with permission from Popular Electronics, 9/91, p. 18. © Copyright Gernsback Publications, Inc., 1991.
Fig. 43-14. Reprinted with permission from Radio-Electronics, 9/87, p. 65. © Copyright Gernsback Publications, Inc., 1987.

Chapter 44

Fig. 44-1. Reprinted with permission from Electronic Design, 5/91, p. 117. Copyright 1991, Penton Publishing.

Fig. 44-2. Reprinted with permission from Electronic Design, 12/89, p. 71. Copyright 1991, Penton Publishing.

Chapter 45

Fig. 45-1. Intersil, Applications Handbook, 1988, p. 5-8.

Fig. 45-2. Electronic Engineering, 4/90, p. 27.

Fig. 45-3. Reprinted from EDN, 10/25/90, p. 79. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 46

Fig. 46-1. Reprinted with permission from Electronic Design, 2/91, p. 81. Copyright 1991, Penton Publishing.

Fig. 46-2. Reprinted with permission from Popular Electronics, 8/89, p. 26-27. © Copyright Gernsback Publications, Inc., 1989.

Fig. 46-3. Reprinted with permission from Radio-Electronics, 4/91, p. 33-42. © Copyright Gernsback Publications, Inc., 1991.

Fig. 46-4. Reprinted with permission from Popular Electronics, 7/91, p. 59-61. © Copyright Gernsback Publications, Inc., 1991.

Fig. 46-5. Reprinted with permission from R-E Experimenter's Handbook, 1989, p. 45. © Copyright Gernsback Publications, Inc., 1989.

Fig. 46-6. Reprinted with permission from Popular Electronics, 10/77, p. 61. © Copyright Gernsback Publications, Inc., 1977.

Fig. 46-7. Reprinted with permission from Popular Electronics, 4/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-8. Reprinted from EDN, 9/3/90, p. 171. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 46-9. Reprinted with permission from Wels' Think Tank, Gernsback Publishing Inc., p. 1.

Fig. 46-10. Reprinted from EDN, 9/90, p. 94. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 46-11. Reprinted with permission from Electronic Design, 5/91, p. 120. Copyright 1991, Penton Publishing.

Fig. 46-12. Reprinted with permission from Radio-Electronics, 8/90, p. 13. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-13. Reprinted with permission from Popular Electronics, 2/90, p. 86. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-14. Reprinted with permission from Electronic Design, 11/82, p. 184. Copyright 1982, Penton Publishing.

Fig. 46-15. Ham Radio, 5/89, p. 85-87.

Fig. 46-16. Reprinted with permission from Popular Electronics, 10/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-17. Reprinted with permission from Popular Electronics, 2/90, p. 23. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-18. Reprinted with permission from Popular Electronics, 8/90, p. 82. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-19. Reprinted with permission from Electronic Design, 4/91, p. 102. Copyright 1991, Penton Publishing.

Fig. 46-20. Reprinted with permission from Popular Electronics, 1/90, p. 22. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-21. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 46. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-22. Reprinted with permission from Electronic Design, 6/91, p. 101. Copyright 1991, Penton Publishing.

Fig. 46-23. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 43-47. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-24. Reprinted with permission from Popular Electronics, 2/90, p. 86. © Copyright Gernsback Publications, Inc., 1990.

Fig. 46-25. Reprinted with permission from Popular Electronics, 7/91, p. 71. © Copyright Gernsback Publications, Inc., 1991.

Fig. 46-26. Reprinted with permission from Popular Electronics, 8/89, p. 26. © Copyright Gernsback Publications, Inc., 1989.

Fig. 46-27. Reprinted with permission from Wels' Think Tank, Gernsback Publishing Inc., p. 1.

Fig. 46-28. Reprinted with permission from Popular Electronics, 10/90, p. 23-24. © Copyright Gernsback Publications, Inc., 1990.

Chapter 47

Fig. 47-1. Reprinted from EDN, Special Ideas Issue, Vol. I, 7/88, p. 39. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 47-2. Reprinted from EDN, 5/91, p. 175. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 47-3. Elektor Electronics, 7-8/91, p. 47.
Fig. 47-4. Hands-On Electronics, 9/87, p. 97.
Fig. 47-5. Reprinted with permission from Popular Electronics, 2/91, p. 25. © Copyright Gernsback Publications, Inc., 1991.
Fig. 47-6. Reprinted with permission from Radio-Electronics, 3/89, p. 73. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-7. Reprinted with permission from Radio-Electronics, 9/89, p. 62. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-8. Reprinted with permission from Popular Electronics/Hands-On Electronics, 2/89, p. 36. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-9. Reprinted with permission from Popular Electronics, 11/91, p. 85. © Copyright Gernsback Publications, Inc., 1991.
Fig. 47-10. Reprinted with permission from Radio-Electronics, 9/89, p. 62. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-11. Reprinted from EDN, 7/91, p. 165. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 47-12. Reprinted with permission from Popular Electronics, 11/91, p. 84. © Copyright Gernsback Publications, Inc., 1991.
Fig. 47-13. Reprinted with permission from Popular Electronics, 3/89, p. 26. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-14. Reprinted with permission from Radio-Electronics, 9/89, p. 62. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-15. Elektor Electronics, 7/89, p. 542.
Fig. 47-16. Reprinted with permission from Popular Electronics, 12/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.
Fig. 47-17. Ham Radio, 3/90, p. 82–83.
Fig. 47-18. Reprinted with permission from Popular Electronics, 1/91, p. 25. © Copyright Gernsback Publications, Inc., 1991.
Fig. 47-19. Reprinted with permission from Radio-Electronics, 9/89, p. 61. © Copyright Gernsback Publications, Inc., 1989.
Fig. 47-20. Reprinted with permission from Popular Electronics, 9/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.

Chapter 48

Fig. 48-1. QST, 7/90, p. 18–19.
Fig. 48-2. 73 Amateur Radio, 2/91, p. 20.
Fig. 48-3. Reprinted with permission from Popular Electronics, 10/89, p. 23. © Copyright Gernsback Publications, Inc., 1989.
Fig. 48-4. Reprinted with permission from Popular Electronics, 7/91, p. 56. © Copyright Gernsback Publications, Inc., 1991.
Fig. 48-5. Reprinted with permission from Radio-Electronics, 12/90, p. 35. © Copyright Gernsback Publications, Inc., 1990.
Fig. 48-6. Reprinted with permission from Popular Electronics, 11/90, p. 70. © Copyright Gernsback Publications, Inc., 1990.
Fig. 48-7. Ham Radio, 6/90, p. 44–47.
Fig. 48-8. Reprinted with permission from Popular Electronics, 7/91, p. 58. © Copyright Gernsback Publications, Inc., 1991.
Fig. 48-9. Reprinted with permission from Popular Electronics, 11/90, p. 70. © Copyright Gernsback Publications, Inc., 1990.
Fig. 48-10. Reprinted with permission from Popular Electronics, 7/91, p. 57. © Copyright Gernsback Publications, Inc., 1991.
Fig. 48-11. 73 Amateur Radio, 3/89, p. 66.

Chapter 49

Fig. 49-1. Reprinted with permission from Popular Electronics, 10/89, p. 84. © Copyright Gernsback Publications, Inc., 1989.
Fig. 49-2. Reprinted with permission from Radio-Electronics, 5/85, p.73–77. © Copyright Gernsback Publications, Inc., 1985.
Fig. 49-3. Elektor Electronics, 7/89, p. S35.
Fig. 49-4. Elektor Electronics, 7/89, p. S32.
Fig. 49-5. Reprinted with permission from Popular Electronics, 10/90, p. 26. © Copyright Gernsback Publications, Inc., 1990.
Fig. 49-6. Reprinted with permission from Radio-Electronics, 1/91, p. 71. © Copyright Gernsback Publications, Inc., 1991.
Fig. 49-7. Reprinted with permission from Radio-Electronics, 1/91, p. 71. © Copyright Gernsback Publications, Inc., 1991.
Fig. 49-8. Reprinted with permission from Popular Electronics, 9/90, p. 82. © Copyright Gernsback Publications, Inc., 1990.
Fig. 49-9. Reprinted with permission from Popular Elec-

tronics, 9/90, p. 83. © Copyright Gernsback Publications, Inc., 1990.

Chapter 50

Fig. 50-1. Reprinted with permission from Electronic Design, 11/88, p. 144. Copyright 1988, Penton Publishing.

Fig. 50-2. Reprinted with permission from Popular Electronics, 7/89, p. 26. © Copyright Gernsback Publications, Inc., 1989.

Fig. 50-3. Reprinted with permission from Popular Electronics, 9/90, p. 22. © Copyright Gernsback Publications, Inc., 1990.

Fig. 50-4. Reprinted with permission from Popular Electronics/Hands-On Electronics, 12/88, p. 103. © Copyright Gernsback Publications, Inc., 1988.

Fig. 50-5. Reprinted with permission from Popular Electronics, 7/89, p. 26. © Copyright Gernsback Publications, Inc., 1989.

Chapter 51

Fig. 51-1. QST, 5/89, p. 31.

Fig. 51-2. QST, 5/89, p. 31.

Fig. 51-3. QST, 5/89, p. 32.

Fig. 51-4. QST, 5/89, p. 32.

Fig. 51-5. QST, 5/89, p. 33.

Fig. 51-6. QST, 5/89, p. 32.

Chapter 52

Fig. 52-1. Reprinted from EDN, 6/91, p. 161. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 52-2. Reprinted with permission from Electronic Design, 2/91, p. 102. Copyright 1991, Penton Publishing.

Fig. 52-3. ARRL Handbook, 1991, p. 18 – 7.

Fig. 52-4. Reprinted with permission from Popular Electronics, 5/91, p. 25 – 26. © Copyright Gernsback Publications, Inc., 1991.

Fig. 52-5. Ham Radio, 4/89, p. 65.

Fig. 52-6. Reprinted with permission from Electronic Design, 11/90, p. 136. Copyright 1990, Penton Publishing.

Fig. 52-7. Reprinted with permission from Radio-Electronics, 9/89, p. 59. © Copyright Gernsback Publications, Inc., 1989.

Fig. 52-8. Reprinted with permission from Popular Electronics, 2/91, p. 28. © Copyright Gernsback Publications, Inc., 1991.

Fig. 52-9. 73 Amateur Radio, 7/90, p. 70.

Fig. 52-10. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.

Fig. 52-11. Reprinted from EDN, 6/91, p. 182. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 52-12. Maxim, Seminar Applications Book, 1988/89, p. 79.

Fig. 52-13. Reprinted with permission from Radio-Electronics, 9/89, p. 60. © Copyright Gernsback Publications, Inc., 1989.

Fig. 52-14. Electronics Today International, 4/78, p. 31.

Fig. 52-15. Reprinted with permission from Popular Electronics, 9/90, p. 22. © Copyright Gernsback Publications, Inc., 1990.

Chapter 53

Fig. 53-1. Reprinted from EDN, 9/90, p. 172. © 1990 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 53-2. Reprinted with permission from Popular Electronics, 3/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.

Fig. 53-3. Elektor Electronics, 7/89, p. S38.

Fig. 53-4. Reprinted with permission from Electronic Design, 7/82, p. 194. Copyright 1991, Penton Publishing.

Fig. 53-5. Reprinted with permission from Radio-Electronics, 2/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.

Fig. 53-6. Reprinted with permission from Popular Electronics, 12/77, p. 83. © Copyright Gernsback Publications, Inc., 1977.

Fig. 53-7. ARRL Handbook, 1991, p. 12 – 17.

Fig. 53-8. Reprinted with permission from Radio-Electronics, 4/70, p. 29. © Copyright Gernsback Publications, Inc., 1970.

Chapter 54

Fig. 54-1. Reprinted with permission from Popular Electronics, 12/90, p. 22. © Copyright Gernsback Publications, Inc., 1990.

Fig. 54-2. Reprinted with permission from Popular Electronics, 3/90, p. 70. © Copyright Gernsback Publications, Inc., 1990.

Fig. 54-3. Elektor Electronics, 7/89 Supplement, 545.

Chapter 55

Fig. 55-1. Elektor Electronics, 12/90, p. 54.

Fig. 55-2. Reprinted with permission from Popular Elec-

tronics, 9/90, p. 84. © Copyright Gernsback Publications, Inc., 1990.
Fig. 55-3. RF Design, 7/88, p. 43.
Fig. 55-4. Reprinted with permission from Popular Electronics, 11/90, p. 75. © Copyright Gernsback Publications, Inc., 1990.
Fig. 55-5. Reprinted with permission from Radio-Electronics, 7/90, p. 59. © Copyright Gernsback Publications, Inc., 1990.
Fig. 55-6. Reprinted with permission from Popular Electronics, 5/91, p. 76. © Copyright Gernsback Publications, Inc., 1991.

Chapter 56

Fig. 56-1. Elektor Electronics, 7/89 Supplement, p. 39–40.
Fig. 56-2. Reprinted from EDN, Design Ideas Special Issue, 7/20/89, Vol. IV, p. 36. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 56-3. Reprinted from EDN, 12/89, p. 260. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 56-4. Reprinted from EDN, 12/89, p. 257. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 56-5. Reprinted from EDN, 5/91, p. 121. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 56-6. Reprinted with permission from Popular Electronics, 8/91, p. 35–36. © Copyright Gernsback Publications, Inc., 1991.
Fig. 56-7. Harris, Analog Product Data Book, 1988, p. 10–162.
Fig. 56-8. Reprinted with permission from Radio-Electronics, 1/91, p. 9–10. © Copyright Gernsback Publications, Inc., 1991.

Chapter 57

Fig. 57-1. ARRL Handbook, 1991, p. 12–31.
Fig. 57-2. ARRL Handbook, 1991, p. 12–31.
Fig. 57-3. ARRL Handbook, 1991, p. 12–31.

Chapter 58

Fig. 58-1. Reprinted from EDN, 2/91, p. 183. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 58-2. Reprinted with permission from Electronic Design, 12/90, p. 63. Copyright 1990, Penton Publishing.
Fig. 58-3. Reprinted from EDN, 7/91, p. 165. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 58-4. Reprinted with permission from Popular Electronics, 11/90, p. 76. © Copyright Gernsback Publications, Inc., 1990.
Fig. 58-5. Reprinted with permission from Radio-Electronics, 1/89, p. 69. © Copyright Gernsback Publications, Inc., 1989.
Fig. 58-6. Reprinted with permission from Electronic Design, 5/91, p. 121. Copyright 1991, Penton Publishing.
Fig. 58-7. Reprinted with permission from Electronic Design, 12/88, p. 69. Copyright 1988, Penton Publishing.
Fig. 58-8. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.
Fig. 58-9. Maxim, Seminar Applications Book, 1988/89, p. 58.
Fig. 58-10. Reprinted with permission from Radio-Electronics, 1/89, p. 69. © Copyright Gernsback Publications, Inc., 1989.

Chapter 59

Fig. 59-1. Reprinted from EDN, 1/90, p. 138. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 59-2. Elektor Electronics, 6/91, p. 43–45.
Fig. 59-3. Ham Radio, 4/90, p. 169.
Fig. 59-4. Elektor Electronics, 6/91, p. 43–45.
Fig. 59-5. QST, 10/90, p. 25.
Fig. 59-6. Hands-On Electronics, 11/87, p. 93.
Fig. 59-7. Reprinted with permission from Popular Electronics, 1/91, p. 27. © Copyright Gernsback Publications, Inc., 1991.
Fig. 59-8. Reprinted with permission from Popular Electronics, 10/89, p. 22. © Copyright Gernsback Publications, Inc., 1989.

Chapter 60

Fig. 60-1. Reprinted with permission from Popular Electronics, 7/91, p. 71–72. © Copyright Gernsback Publications, Inc., 1991.
Fig. 60-2. Reprinted from EDN, 11/90, p. 234. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 60-3. QST, 6/91, p. 27–29.
Fig. 60-4. Reprinted with permission from Wels' Think Tank, Gernsback Publishing Inc., p. 37.
Fig. 60-5. Reprinted with permission from Electronic

Design, 12/88, p. 103. Copyright 1988, Penton Publishing.

Fig. 60-6. Reprinted with permission from Popular Electronics/Hands-On Electronics, 5/89, p. 86. © Copyright Gernsback Publications, Inc., 1989.

Fig. 60-7. Reprinted with permission from Hands-On Electronics/Popular Electronics, 1/89, p. 26.

Fig. 60-8. Reprinted with permission from Radio-Electronics, 8/86, p. 83. © Copyright Gernsback Publications, Inc., 1986.

Fig. 60-9. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 2.

Fig. 60-10. Reprinted with permission from Radio-Electronics, 1/89, p. 71. © Copyright Gernsback Publications, Inc., 1989.

Fig. 60-11. Harris, Analog Product Data Book, 1988, p. 10-147.

Chapter 61

Fig. 61-1. Reprinted with permission from Electronic Design, 9/89, p. 93–94. Copyright 1989, Penton Publishing.

Fig. 61-2. Reprinted with permission from Popular Electronics, Fact Card No. 65. © Copyright Gernsback Publications, Inc.

Fig. 61-3. Reprinted with permission from Popular Electronics, 3/91, p. 40–42. © Copyright Gernsback Publications, Inc., 1991.

Fig. 61-4. Reprinted with permission from Popular Electronics, 1/91, p. 28–29. © Copyright Gernsback Publications, Inc., 1991.

Fig. 61-5. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 39.

Fig. 61-6. Reprinted with permission from Popular Electronics, Fact Card No. 170. © Copyright Gernsback Publications, Inc.

Fig. 61-7. Reprinted with permission from Popular Electronics/Hands-On Electronics, 12/88, p. 58. © Copyright Gernsback Publications, Inc., 1988.

Chapter 62

Fig. 62-1. Reprinted from EDN, 1/91, p. 154–156. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 62-2. Hands-On Electronics, Fact Card No. 57.

Fig. 62-3. Reprinted from EDN, Design Ideas Special Issue, Vol. IV, 7/20/89, p. 12. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 62-4. Reprinted from EDN, 2/91, p. 106. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 62-5. Reprinted with permission from Popular Electronics, Fact Card No. 65. © Copyright Gernsback Publications, Inc.

Fig. 62-6. Reprinted with permission from Popular Electronics, 8/90, p. 20. © Copyright Gernsback Publications, Inc., 1990.

Fig. 62-7. Reprinted with permission from Radio-Electronics, 9/87, p. 64. © Copyright Gernsback Publications, Inc., 1987.

Fig. 62-8. Reprinted with permission from Electronic Design, 6/22/89, p. 105. Copyright 1989, Penton Publishing.

Fig. 62-9. Hands-On Electronics, Fact Card No. 40.

Chapter 63

Fig. 63-1. Reprinted from EDN, Design Ideas Special Issue, Vol. IV, 7/20/89, p. 14. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 63-2. Reprinted from EDN, 1/90, p. 218. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 63-3. Electronic Engineering, 9/88, p. 31.

Fig. 63-4. Reprinted from EDN, 6/91, p. 162. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 63-5. Reprinted with permission from Electronic Design, 11/90, p. 136. Copyright 1990, Penton Publishing.

Fig. 63-6. Reprinted with permission from Popular Electronics, 6/89, p. 25. © Copyright Gernsback Publications, Inc., 1989.

Fig. 63-7. Reprinted with permission from Electronic Design, 4/91, p. 104. Copyright 1991, Penton Publishing.

Fig. 63-8. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 52.

Fig. 63-9. Elektor Electronics, 12/90 Supplement, p. 41.

Fig. 63-10. Reprinted with permission from Radio-Electronics, 10/90, p. 14. © Copyright Gernsback Publications, Inc., 1990.

Fig. 63-11. Elektor Electronics, 12/90 Supplement, p. 536.

Fig. 63-12. Reprinted with permission from Radio-Electronics, 3/89, p. 53. © Copyright Gernsback Publications, Inc., 1989.

Fig. 63-13. Reprinted with permission from Electronic Design, 2/91, p. 104. Copyright 1991, Penton Publishing.

Fig. 63-14. Reprinted with permission from Electronic Design, 3/91, p. 86. Copyright 1991, Penton Publishing.
Fig. 63-15. 73 Amateur Radio, 11/90, p. 67.
Fig. 63-16. 73 Amateur Radio, 7/90, p. 68.
Fig. 63-17. 73 Amateur Radio, 7/90, p. 69.
Fig. 63-18. 73 Amateur Radio, 7/90, p. 77.
Fig. 63-19. Reprinted with permission from Electronic Design, 12/90, p. 77. Copyright 1990, Penton Publishing.
Fig. 63-20. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.
Fig. 63-21. Linear Technology, Design Note #47.
Fig. 63-22. Linear Technology, Design Note #47.
Fig. 63-23. Reprinted with permission from Radio-Electronics, 5/91, p. 49–55. © Copyright Gernsback Publications, Inc., 1991.
Fig. 63-24. Elektor Electronics, 7-8/89, p. 28.
Fig. 63-25. Ham Radio, 4/89, p. 67.
Fig. 63-26. Linear Technology, Linear Databook, 1986, p. 2–112.
Fig. 63-27. Reprinted from EDN, 1/91, p. 152. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 63-28. Reprinted from EDN, 8/2/90, p. 96. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 63-29. Reprinted with permission from Radio-Electronics, 3/89, p. 54–55. © Copyright Gernsback Publications, Inc., 1989.
Fig. 63-30. Reprinted with permission from Electronic Design, 6/91, p. 109. Copyright 1991, Penton Publishing.
Fig. 63-31. Reprinted with permission from Electronic Design, 3/91, p. 117. Copyright 1991, Penton Publishing.
Fig. 63-32. Reprinted with permission from Popular Electronics, Fact Card No. 134. © Copyright Gernsback Publications, Inc.

Chapter 64

Fig. 64-1. Reprinted with permission from Radio-Electronics, 8/91, p. 42. © Copyright Gernsback Publications, Inc., 1991.
Fig. 64-2. Linear Technology, Application Note 45.
Fig. 64-3. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 61. © Copyright Gernsback Publications, Inc., 1990.
Fig. 64-4. Reprinted with permission from Popular Electronics, 6/89, p. 88. © Copyright Gernsback Publications, Inc., 1989.
Fig. 64-5. Reprinted with permission from Radio-Electronics, 2/91, p. 34. © Copyright Gernsback Publications, Inc., 1991.

Chapter 65

Fig. 65-1. QST, 9/89, p. 23.
Fig. 65-2. Reprinted with permission from Radio-Electronics, 3/90, p. 31–34. © Copyright Gernsback Publications, Inc., 1990.
Fig. 65-3. Reprinted with permission from Popular Electronics, 12/90, p. 22. © Copyright Gernsback Publications, Inc., 1990.
Fig. 65-4. Hands-On Electronics, Fact Card No. 40.
Fig. 65-5. Reprinted with permission from Popular Electronics, 9/90, p. 27. © Copyright Gernsback Publications, Inc., 1990.
Fig. 65-6. Reprinted with permission from Popular Electronics, 4/90, p. 45–46. © Copyright Gernsback Publications, Inc., 1990.
Fig. 65-7. Linear Technology, 10/90, Design Notes #40.
Fig. 65-8. Reprinted with permission from Popular Electronics, 11/90, p. 24. © Copyright Gernsback Publications, Inc., 1990.
Fig. 65-9. Reprinted with permission from Radio-Electronics, 1/89, p. 70. © Copyright Gernsback Publications, Inc., 1989.

Chapter 66

Fig. 66-1. Reprinted from EDN, 11/90, p. 292. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 66-2. Reprinted with permission from Popular Electronics, 6/89, p. 26. © Copyright Gernsback Publications, Inc., 1989.
Fig. 66-3. Reprinted with permission from Electronic Design, 5/90, p. 80. Copyright 1990, Penton Publishing.
Fig. 66-4. Ham Radio, 4/90, p. 73–78.
Fig. 66-5. Reprinted from EDN, 1/91, p. 170. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 66-6. Reprinted with permission from Electronic
Fig. 66-8. Hands-On Electronics, Fact Card No. 37.
Fig. 66-9. Reprinted from EDN, 4/91, p. 188. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Design, 9/89, p. 94–95. Copyright 1989, Penton Publishing.

Fig. 66-7. Reprinted from EDN, 12/90, p. 218. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 67

Fig. 67-1. Reprinted with permission from Popular Electronics, 10/89, p. 104. © Copyright Gernsback Publications, Inc., 1989.

Fig. 67-2. Elektor Electronics, 12/90 Supplement, p. 42.

Fig. 67-3. Elektor Electronics, 7-8/89, p. 67.

Fig. 67-4. Reprinted with permission from Popular Electronics, 3/91, p. 55–57. © Copyright Gernsback Publications, Inc., 1991.

Fig. 67-5. Reprinted with permission from Popular Electronics, 7/91, p. 57. © Copyright Gernsback Publications, Inc., 1991.

Fig. 67-6. Reprinted with permission from Radio-Electronics, 12/75, p. 42. © Copyright Gernsback Publications, Inc., 1975.

Fig. 67-7. Reprinted with permission from Popular Electronics/Hands-On Electronics, 5/89, p. 28. © Copyright Gernsback Publications, Inc., 1989.

Fig. 67-8. TAB Books, Third Book of Electronic Projects, p. 39.

Chapter 68

Fig. 68-1. Reprinted with permission from Radio-Electronics, 5/91, p. 40–42. © Copyright Gernsback Publications, Inc., 1991.

Fig. 68-2. Reprinted with permission from Popular Electronics, 8/90, p. 80. © Copyright Gernsback Publications, Inc., 1990.

Fig. 68-3. GE, Application Note 90.16, p. 29.

Fig. 68-4. Reprinted with permission from Popular Electronics, 5/91, p. 74. © Copyright Gernsback Publications, Inc., 1991.

Fig. 68-5. Reprinted with permission from Radio-Electronics, 12/88, p. 76. © Copyright Gernsback Publications, Inc., 1988.

Fig. 68-6. Elektor Electronics, 302 Circuits, p. 203.

Fig. 68-7. Reprinted with permission from Popular Electronics, 8/90, p. 80. © Copyright Gernsback Publications, Inc., 1990.

Fig. 68-8. Elektor Electronics, 302 Circuits, p. 165–166.

Fig. 68-9. Reprinted from EDN, 3/75, p. 73. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 68-10. Reprinted from EDN, 10/90, p. 238. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 68-11. Hands-On Electronics, Fact Card No. 86.

Chapter 69

Fig. 69-1. Reprinted with permission from R-E Experimenter's Handbook, 1987, p. 122. © Copyright Gernsback Publications, Inc., 1987.

Fig. 69-2. Reprinted with permission from R-E Experimenter's Handbook, 1987, p. 122. © Copyright Gernsback Publications, Inc., 1987.

Chapter 70

Fig. 70-1. Reprinted with permission from Popular Electronics, 6/89, p. 59–62. © Copyright Gernsback Publications, Inc., 1989.

Fig. 70-2. Reprinted with permission from Radio-Electronics, 12/90, p. 81. © Copyright Gernsback Publications, Inc., 1990.

Fig. 70-3. Reprinted from EDN, 3/79, p. 130. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 70-4. Reprinted with permission from Radio-Electronics, 12/90, p. 82. © Copyright Gernsback Publications, Inc., 1990.

Fig. 70-5. Reprinted with permission from Popular Electronics, 11/90, p. 77. © Copyright Gernsback Publications, Inc., 1990.

Chapter 71

Fig. 71-1. 73 Amateur Radio, 11/90, p. 34.

Fig. 71-2. 73 Amateur Radio, 4/91, p. 26.

Fig. 71-3. Reprinted with permission from Popular Electronics, 8/91, p. 72. © Copyright Gernsback Publications, Inc., 1991.

Fig. 71-4. Reprinted with permission from Popular Electronics, 5/90, p. 85 and 96. © Copyright Gernsback Publications, Inc., 1990.

Fig. 71-5. Reprinted with permission from Popular Electronics, 10/89, p. 24. © Copyright Gernsback Publications, Inc., 1989.

Fig. 71-6. Reprinted with permission from Popular Electronics, 3/90, p. 96. © Copyright Gernsback Publications, Inc., 1990.

Fig. 71-7. Reprinted with permission from Popular Electronics, 2/82, p. 99. © Copyright Gernsback Publications, Inc., 1982.

Fig. 71-8. Reprinted with permission from Radio-Elec-

tronics, 1/91, p. 56. © Copyright Gernsback Publications, Inc., 1991.
Fig. 71-9. Reprinted with permission from Popular Electronics, 10/89, p. 22. © Copyright Gernsback Publications, Inc., 1989.

Chapter 72

Fig. 72-1. QST, 8/90, p. 59.
Fig. 72-2. ARRL Handbook, 1991, p. 12-30.
Fig. 72-3. Reprinted with permission from Popular Electronics, 3/90, p. 48. © Copyright Gernsback Publications, Inc., 1990.
Fig. 72-4. ARRL Handbook, 1991, p. 12-26.
Fig. 72-5. 73 Amateur Radio, 3/90, p. 13.
Fig. 72-6. ARRL Handbook, 1991, p. 14-6.

Chapter 73

Fig. 73-1. Elektor Electronics, 7/89 Supplement, p. S33.
Fig. 73-2. Reprinted from EDN, Design Ideas Special Issue, Vol. IV, 7/20/89, p. 28-29. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 73-3. Reprinted with permission from Electronic Design, 12/88, p. 105. Copyright 1988, Penton Publishing.
Fig. 73-4. Reprinted from EDN, 4/91, p. 179. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 73-5. Reprinted with permission from Electronic Design, 2/91, p. 104. Copyright 1991, Penton Publishing.
Fig. 73-6. Reprinted from EDN, 6/90, p. 266. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 73-7. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.

Chapter 74

Fig. 74-1. Elektor Electronics, 12/90, p. 512-513.
Fig. 74-2. Intersil, Applications Handbook, 1988, p. 6-20.
Fig. 74-3. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.

Chapter 75

Fig. 75-1. GE, Optoelectronics, Third Edition, p. 139.
Fig. 75-2. Reprinted from EDN, 12/89, p. 258. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 75-3. Elektor Electronics, 12/90, p. 525.
Fig. 75-4. Reprinted with permission from Popular Electronics, 5/91, p. 86. © Copyright Gernsback Publications, Inc., 1991.
Fig. 75-5. Reprinted from EDN, 7/90, p. 185. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 75-6. Reprinted from EDN, 9/89, p. 170. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 76

Fig. 76-1. QST, 7/90, p. 29-31.
Fig. 76-2. CQ, 7/90, p. 61-65.
Fig. 76-3. Elektor Electronics, 301 Circuits, p. 38-39.
Fig. 76-4. QST, 10/90, p. 18-21.
Fig. 76-5. North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804.
Fig. 76-6. Elektor Electronics, 12/90, p. 57-58.
Fig. 76-7. Reprinted with permission from Popular Electronics, 1/91, p. 58. © Copyright Gernsback Publications, Inc., 1991.
Fig. 76-8. QST, 7/90, p. 28-30.
Fig. 76-9. Reprinted with permission from Popular Electronics, 1/91, p. 58. © Copyright Gernsback Publications, Inc., 1991.
Fig. 76-10. Ham Radio, 11/88, p. 60-62.
Fig. 76-11. Reprinted with permission from Popular Electronics, 12/89, p. 23. © Copyright Gernsback Publications, Inc., 1989.
Fig. 76-12. Reprinted with permission from Popular Electronics, 1/91, p. 59. © Copyright Gernsback Publications, Inc., 1991.
Fig. 76-13. Reprinted with permission from Popular Electronics, 1/91, p. 59. © Copyright Gernsback Publications, Inc., 1991.
Fig. 76-14. Reprinted with permission from Popular Electronics, 1/91, p. 59. © Copyright Gernsback Publications, Inc., 1991.
Fig. 76-15. Reprinted with permission from Popular Electronics, 1/90, p. 98. © Copyright Gernsback Publications, Inc., 1990.
Fig. 76-16. Reprinted with permission from Popular Electronics, 11/90, p. 69. © Copyright Gernsback Publications, Inc., 1990.
Fig. 76-17. Reprinted with permission from Popular Electronics, 1/90, p. 98. © Copyright Gernsback Publications, Inc., 1990.

Fig. 76-18. Reprinted with permission from Popular Electronics, 1/90, p. 74. © Copyright Gernsback Publications, Inc., 1990.
Fig. 76-19. Copyright of Motorola, MOSPOWER Design Catalog, 1/83, p. 5–6. Used by permission.
Fig. 76-20. Reprinted with permission from Popular Electronics, 1/90, p. 74. © Copyright Gernsback Publications, Inc., 1990.
Fig. 76-21. Reprinted with permission from Popular Electronics, 1/90, p. 74. © Copyright Gernsback Publications, Inc., 1990.
Fig. 76-22. Reprinted with permission from Popular Electronics, 1/90, p. 74. © Copyright Gernsback Publications, Inc., 1990.

Chapter 77

Fig. 77-1. Elektor Electronics, 7/89, p. S31.
Fig. 77-2. North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804.
Fig. 77-3. Reprinted with permission from Popular Electronics, 11/90, p. 85. © Copyright Gernsback Publications, Inc., 1990.
Fig. 77-4. Elektor Electronics, 3/91, p. 37–41.
Fig. 77-5. Reprinted with permission from Popular Electronics, 3/90, p. 48. © Copyright Gernsback Publications, Inc., 1990.
Fig. 77-6. Reprinted with permission from Popular Electronics, 8/90, p. 88. © Copyright Gernsback Publications, Inc., 1990.
Fig. 77-7. Reprinted with permission from Radio-Electronics, 10/89, p. 43. © Copyright Gernsback Publications, Inc., 1989.
Fig. 77-8. Ham Radio, 4/90, p. 21.
Fig. 77-9. Reprinted with permission from Radio-Electronics, 11/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.

Chapter 78

Fig. 78-1. Reprinted from EDN, 12/90, p. 153. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 79

Fig. 79-1. Elektor Electronics, 12/90, p. 523.
Fig. 79-2. Reprinted from EDN, 10/90, p. 206. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 79-3. Elektor Electronics, 7/89, p. 542.
Fig. 79-4. Micro Linear Advertisement
Fig. 79-5. Reprinted with permission from Popular Electronics, 3/91, p. 92. © Copyright Gernsback Publications, Inc., 1991.
Fig. 79-6. Reprinted with permission from Popular Electronics, 1/90, p. 85. © Copyright Gernsback Publications, Inc., 1990.
Fig. 79-7. Reprinted with permission from Electronic Design, 4/91, p. 99. Copyright 1991, Penton Publishing.
Fig. 79-8. Reprinted with permission from Popular Electronics, 1/90, p. 84. © Copyright Gernsback Publications, Inc., 1990.
Fig. 79-9. Reprinted with permission from Radio-Electronics, 7/89, p. 52. © Copyright Gernsback Publications, Inc., 1989.
Fig. 79-10. Reprinted with permission from Popular Electronics, 1/90, p. 101. © Copyright Gernsback Publications, Inc., 1990.
Fig. 79-11. Reprinted from EDN, 11/90, p. 236. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 79-12. GE/RCA, BiMOS Operational Amplifier Circuit Ideas, 1987, p. 7.
Fig. 79-13. 73 Amateur Radio, 12/76, p. 97–99.
Fig. 79-14. Reprinted with permission from Popular Electronics, 3/91, p. 88. © Copyright Gernsback Publications, Inc., 1991.
Fig. 79-15. Reprinted from EDN, 11/90, p. 23. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 80

Fig. 80-1. Hands-On Electronics, 1-2/86, p. 67.
Fig. 80-2. Reprinted with permission from Radio-Electronics, 7/90, p. 58. © Copyright Gernsback Publications, Inc., 1990.
Fig. 80-3. Reprinted with permission from Radio-Electronics, 7/90, p. 57. © Copyright Gernsback Publications, Inc., 1990.
Fig. 80-4. Reprinted with permission from National Semiconductor Corp., CMOS Databook, 1981, p. 8–46.
Fig. 80-5. Reprinted with permission from Popular Electronics, 7/89, p. 22. © Copyright Gernsback Publications, Inc., 1989.
Fig. 80-6. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.

Chapter 81

Fig. 81-1. Elektor Electronics, 7/89 Supplement, p. 41–42.
Fig. 81-2. Elektor Electronics, 7/89, p. 537.
Fig. 81-3. Reprinted with permission from Popular Electronics, 10/89, p. 85. © Copyright Gernsback Publications, Inc., 1989.
Fig. 81-4. Reprinted with permission from Popular Electronics, 1/91, p. 27. © Copyright Gernsback Publications, Inc., 1991.
Fig. 81-5. Reprinted with permission from Popular Electronics, 9/90, p. 63–66. © Copyright Gernsback Publications, Inc., 1990.
Fig. 81-6. Elektor Electronics, 301 Circuits, p. 130.
Fig. 81-7. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.
Fig. 81-8. Reprinted from EDN, 8/2/91, p. 112. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 82

Fig. 82-1. Reprinted with permission from Radio-Electronics, 4/87, p. 30. © Copyright Gernsback Publications, Inc., 1987.
Fig. 82-2. Hands-On Electronics, 11/86, p. 32.
Fig. 82-3. Reprinted from EDN, 6/88, p. 175. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 82-4. Reprinted with permission from Electronic Design, 10/89, p. 85. Copyright 1991, Penton Publishing.

Chapter 83

Fig. 83-1. Elektor Electronics, 301 Circuits, p. 32–35.
Fig. 83-2. Reprinted from EDN, 11/90, p. 232. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 83-3. Reprinted by permission of Texas Instruments, Linear and Interface Circuit Applications, 1985, p. 7–15.
Fig. 83-4. Reprinted with permission from Radio-Electronics, 12/88, p. 76. © Copyright Gernsback Publications, Inc., 1988.
Fig. 83-5. Reprinted with permission from Radio-Electronics, 7/89, p. 53. © Copyright Gernsback Publications, Inc., 1989.
Fig. 83-6. Reprinted with permission from Popular Electronics, Fact Card No. 98. © Copyright Gernsback Publications, Inc.
Fig. 83-7. Reprinted with permission from Radio-Electronics, 7/89, p. 53. © Copyright Gernsback Publications, Inc., 1989.
Fig. 83-8. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.
Fig. 83-9. Reprinted by permission of Texas Instruments, Linear and Interface Circuit Applications, 1985, Vol. 1, p. 3–16.
Fig. 83-10. Reprinted with permission from Radio-Electronics, 12/88, p. 76. © Copyright Gernsback Publications, Inc., 1988.

Chapter 84

Fig. 84-1. Elektor Electronics, 12/90, p. 531–532.
Fig. 84-2. Reprinted from EDN, 7/85, p. 282. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 84-3. Reprinted with permission from Popular Electronics, 12/89, p. 29–34. © Copyright Gernsback Publications, Inc., 1989.
Fig. 84-4. Reprinted with permission from Radio-Electronics, 12/89, p. 58. © Copyright Gernsback Publications, Inc., 1989.
Fig. 84-5. QST, 6/91, p. 45.
Fig. 84-6. Reprinted with permission from Popular Electronics, 11/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.
Fig. 84-7. Reprinted with permission from Popular Electronics, 11/90, p. 25. © Copyright Gernsback Publications, Inc., 1990.
Fig. 84-8. Reprinted from EDN, 10/90, p. 132. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 84-9. RF Design, 9/90, p. 76.
Fig. 84-10. Elektor Electronics, 12/90, p. 532.
Fig. 84-11. Elektor Electronics, 12/90 Supplement, p. 43.
Fig. 84-12. Electronics Today International, 4/78, p. 31.

Chapter 85

Fig. 85-1. 73 Amateur Radio, 12/90, p. 54.
Fig. 85-2. Hands-On Electronics, 8/87, p. 92.

Chapter 86

Fig. 86-1. Reprinted with permission from Popular Electronics, 6/91, p. 53–57. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-2. Reprinted with permission from Popular Elec-

tronics, 8/90, p. 33–36. © Copyright Gernsback Publications, Inc., 1990.
Fig. 86-3. Reprinted with permission from Popular Electronics, 8/90, p. 29–32. © Copyright Gernsback Publications, Inc., 1990.
Fig. 86-4. Reprinted with permission from Popular Electronics, 6/89, p. 33. © Copyright Gernsback Publications, Inc., 1989.
Fig. 86-5. Elektor Electronics, 7/89 Supplement, p. 526.
Fig. 86-6. Ham Radio, 12/89, p. 32.
Fig. 86-7. Reprinted with permission from Popular Electronics, 1/91, p. 81. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-8. Reprinted with permission from Radio-Electronics, 7/90, p. 8. © Copyright Gernsback Publications, Inc., 1990.
Fig. 86-9. Reprinted from EDN, 12/90, p. 153–154. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 86-10. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 48.
Fig. 86-11. Reprinted with permission from Popular Electronics, 9/91, p. 21. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-12. Hands-On Electronics, 11/87, p. 85.
Fig. 86-13. TAB Books, The Build-It Book of Electronic Projects, p. 24.
Fig. 86-14. Reprinted with permission from Popular Electronics, 5/91, p. 86. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-15. Reprinted with permission from Popular Electronics, 8/90, p. 80. © Copyright Gernsback Publications, Inc., 1990.
Fig. 86-16. Reprinted with permission from Popular Electronics, 7/91, p. 22. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-17. Reprinted with permission from Popular Electronics, Fact Card No. 170. © Copyright Gernsback Publications, Inc.
Fig. 86-18. Hands-On Electronics, Winter 1985, p. 65.
Fig. 86-19. Reprinted with permission from Popular Electronics, 8/90, p. 81. © Copyright Gernsback Publications, Inc., 1990.
Fig. 86-20. Elektor Electronics, 7/89, p. 540.
Fig. 86-21. Reprinted with permission from Popular Electronics, 9/91, p. 22. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-22. Reprinted with permission from Popular Electronics, 2/91, p. 25–26. © Copyright Gernsback Publications, Inc., 1991.
Fig. 86-23. Reprinted with permission from Popular Electronics, 11/90, p. 26. © Copyright Gernsback Publications, Inc., 1990.
Fig. 86-24. Reprinted with permission from General Electric Semiconductor Department, Optoelectronics, Second Edition, p. 118.

Chapter 87

Fig. 87-1. Reprinted from EDN, 9/89, p. 152. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 87-2. GE/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 25.
Fig. 87-3. Reprinted with permission from Popular Electronics, 1/91, p. 22. © Copyright Gernsback Publications, Inc., 1991.

Chapter 88

Fig. 88-1. Elektor Electronics, 12/90, p. 533.
Fig. 88-2. Elektor Electronics, 7/89, p. 528.
Fig. 88-3. Reprinted with permission from Radio-Electronics, 12/88, p. 72. © Copyright Gernsback Publications, Inc., 1988.
Fig. 88-4. Reprinted with permission from Radio-Electronics, 6/91, p. 58. © Copyright Gernsback Publications, Inc., 1991.
Fig. 88-5. Reprinted with permission from Radio-Electronics, 12/88, p. 71. © Copyright Gernsback Publications, Inc., 1988.
Fig. 88-6. Reprinted from EDN, 4/91, p. 180. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 89

Fig. 89-1. Electronic Engineering, 12/90, p. 28.
Fig. 89-2. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 22–23. © Copyright Gernsback Publications, Inc., 1990.
Fig. 89-3. Reprinted with permission from Electronic Design, 4/90, p. 85–88. Copyright 1991, Penton Publishing.
Fig. 89-4. 42 New Ideas, Gernsback Publications Inc., 1984, p. 13.
Fig. 89-5. Linear Technology, 10/90, Design Notes #40.

Chapter 90

Fig. 90-1. Elektor Electronics, 7/89, p. 536–537.
Fig. 90-2. 73 Amateur Radio, 8/90, p. 82.
Fig. 90-3. Reprinted with permission from Popular Electronics, 12/89, p. 82–83. © Copyright Gernsback Publications, Inc., 1989.

Fig. 90-4. Reprinted with permission from Popular Electronics, 12/89, p. 82. © Copyright Gernsback Publications, Inc., 1989.
Fig. 90-5. Reprinted with permission from Popular Electronics, 12/89, p. 27. © Copyright Gernsback Publications, Inc., 1989.
Fig. 90-6. Hands-On Electronics, Fact Card No. 49.
Fig. 90-7. Reprinted from EDN, 8/2/90, p. 86. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 90-8. 73 Amateur Radio, 2/91, p. 78.
Fig. 90-9. Hands-On Electronics, 1-2/86, p. 52.
Fig. 90-10. Copyright of Motorola, Thyristor Device Data, Series A 1985, p. 1 – 6 – 51. Used by permission.
Fig. 90-11. Reprinted with permission from Popular Electronics, 1/91, p. 80. © Copyright Gernsback Publications, Inc., 1991.

Chapter 91

Fig. 91-1. GE/RCA, BiMOS Operational Amplifier Circuit Ideas, 1987, p. 22.
Fig. 91-2. Reprinted with permission from Radio-Electronics, 2/90, p. 58. © Copyright Gernsback Publications, Inc., 1990.
Fig. 91-3. Hands-On Electronics, 8/87, p. 91.

Chapter 92

Fig. 92-1. Elektor Electronics, 7/89, p. 538-539.
Fig. 92-2. Reprinted with permission from Popular Electronics, 2/91, p. 23 – 24. © Copyright Gernsback Publications, Inc., 1991.
Fig. 92-3. GE/RCA, BiMOS Operational Amplifier Circuit Ideas, 1987, p. 28.
Fig. 92-4. Reprinted with permission from Wels' Think Tank, Gernsback Publications Inc., p. 39.
Fig. 92-5. Reprinted with permission from Radio-Electronics, 7/90, p. 58. © Copyright Gernsback Publications, Inc., 1990.
Fig. 92-6. Hands-On Electronics, 9/87, p. 89.

Chapter 93

Fig. 93-1. 73 Amateur Radio, 6/89, p. 65.
Fig. 93-2. 73 Amateur Radio, 11/90, p. 78.
Fig. 93-3. Reprinted with permission from Radio-Electronics, 6-7/89, p. 45 – 50. © Copyright Gernsback Publications, Inc., 1989.
Fig. 93-4. Elektor Electronics, 7/89, p. 521.
Fig. 93-5. 73 Amateur Radio, 4/91, p. 76.
Fig. 93-6. 73 Amateur Radio, 4/90, p. 46 – 47.
Fig. 93-7. 73 Amateur Radio, 7/90, p. 16.

Chapter 94

Fig. 94-1. Reprinted with permission from R-E Experimenter's Handbook, 1987, p. 93 – 94. © Copyright Gernsback Publications, Inc., 1987.
Fig. 94-2. Reprinted with permission from Radio-Electronics, 7/85, p. 61. © Copyright Gernsback Publications, Inc., 1985.
Fig. 94-3. Reprinted with permission from R-E Experimenter's Handbook, 1987, p. 93. © Copyright Gernsback Publications, Inc., 1987.

Chapter 95

Fig. 95-1. Reprinted from EDN, 12/90, p. 224 – 226. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 95-2. Elektor Electronics, 302 Circuits, p. 180.
Fig. 95-3. Elektor Electronics, 6/91, p. 29.
Fig. 95-4. Reprinted with permission from Radio-Electronics, 12/89, p. 81. © Copyright Gernsback Publications, Inc., 1989.
Fig. 95-5. Reprinted from EDN, 2/90, p. 185. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 95-6. Reprinted with permission from Radio-Electronics, 10/90, p. 59 – 60. © Copyright Gernsback Publications, Inc., 1990.
Fig. 95-7. Reprinted with permission from Radio-Electronics, 7/89, p. 41. © Copyright Gernsback Publications, Inc., 1989.
Fig. 95-8. Electronic Products, 4/88, p. 36.
Fig. 95-9. Reprinted from EDN, 12/90, p. 222. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.
Fig. 95-10. Elektor Electronics, 302 Circuits, p. 78 – 79.
Fig. 95-11. Reprinted with permission from Radio-Electronics, 7/91, p. 73. © Copyright Gernsback Publications, Inc., 1991.

Chapter 96

Fig. 96-1. Reprinted with permission from Popular Electronics, 3/90, p. 28. © Copyright Gernsback Publications, Inc., 1990.
Fig. 96-2. Reprinted from EDN, 3/91, p. 137 – 138. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 97

Fig. 97-1. Elektor Electronics, 7/89 Supplement, p. 34.

Fig. 97-2. Reprinted with permission from Popular Electronics, 6/91, p. 39–40. © Copyright Gernsback Publications, Inc., 1991.

Chapter 98

Fig. 98-1. Elektor Electronics, 12/90 Supplement, p. 41–42.

Fig. 98-2. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.

Fig. 98-3. Reprinted with permission from Popular Electronics, 10/89, p. 105. © Copyright Gernsback Publications, Inc., 1989.

Fig. 98-4. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.

Fig. 98-5. Reprinted from EDN, 10/90, p. 222. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 98-6. Reprinted with permission from Popular Electronics, 11/90, p. 102. © Copyright Gernsback Publications, Inc., 1990

Fig. 98-7. Reprinted from EDN, 5/90, p. 174. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 98-8. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.

Fig. 98-9. Reprinted with permission from Radio-Electronics, 12/89, p. 56. © Copyright Gernsback Publications, Inc., 1989.

Fig. 98-10. RF Design, 4/86, p. 41.

Fig. 98-11. Fairchild Camera and Instrument Corp., Linear Databook, 1982, p. 5–24.

Chapter 99

Fig. 99-1. Reprinted from EDN, 6/91, p. 173. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 99-2. Reprinted with permission from Radio-Electronics, 8/91, p. 63. © Copyright Gernsback Publications, Inc., 1991.

Fig. 99-3. Reprinted with permission from Popular Electronics, 10/89, p. 37–38. © Copyright Gernsback Publications, Inc., 1989.

Fig. 99-4. Reprinted with permission from Radio-Electronics, 1/91, p. 41–43. © Copyright Gernsback Publications, Inc., 1991.

Fig. 99-5. Reprinted with permission from Radio-Electronics, 8/91, p. 62. © Copyright Gernsback Publications, Inc., 1991.

Fig. 99-6. Reprinted with permission from Radio-Electronics, 8/91, p. 63. © Copyright Gernsback Publications, Inc., 1991.

Fig. 99-7. Reprinted with permission from Radio-Electronics, 3/89, p. 33–37. © Copyright Gernsback Publications, Inc., 1989.

Fig. 99-8. Reprinted with permission from R-E Experimenter's Handbook, 1990, p. 45. © Copyright Gernsback Publications, Inc., 1990.

Fig. 99-9. Reprinted from EDN, Design Ideas Special Issue, Vol. IV, 7/20/89, p. 16. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 100

Fig. 100-1. Reprinted with permission from Radio-Electronics, 6/91, p. 55. © Copyright Gernsback Publications, Inc., 1991.

Fig. 100-2. Reprinted from EDN, 5/91, p. 178. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 100-3. Reprinted with permission from Radio-Electronics, 6/91, p. 59–60. © Copyright Gernsback Publications, Inc., 1991.

Fig. 100-4. Reprinted with permission from Radio-Electronics, 6/91, p. 56. © Copyright Gernsback Publications, Inc., 1991.

Chapter 101

Fig. 101-1. Elektor Electronics, 12/90 Supplement, p. 40.

Chapter 102

Fig. 102-1. Reprinted from EDN, 6/91, p. 184. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 102-2. Reprinted from EDN, 7/91, p. 166–168. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Fig. 102-3. Reprinted with permission from Popular Electronics, 11/89, p. 85. © Copyright Gernsback Publications, Inc., 1989.

Fig. 102-4. Reprinted with permission from Electronic Design, 4/89, p. 77. Copyright 1989, Penton Publishing.

Fig. 102-5. Reprinted with permission from Electronic Design, 11/88, p. 78. Copyright 1988, Penton Publishing.

Fig. 102-6. Reprinted with permission from Electronic Design, 3/91, p. 119 – 120. Copyright 1991, Penton Publishing.

Fig. 102-7. Reprinted from EDN, 1/91, p. 168. © 1991 Cahners Publishing Company, a division of Reed Publishing USA.

Chapter 103

Fig. 103-1. Reprinted with permission from R-E Experimeter's Handbook, 1990, p. 155. © Copyright Gernsback Publications, Inc., 1990.

Fig. 103-2. Reprinted with permission from Popular Electronics, 8/89, p. 27. © Copyright Gernsback Publications, Inc., 1989.

Chapter 104

Fig. 104-1. Reprinted with permission from Electronic Design, 7/91, p. 136. Copyright 1991, Penton Publishing.

Fig. 104-2. Reprinted with permission from Popular Electronics, 1/90, p. 26. © Copyright Gernsback Publications, Inc., 1990.

Fig. 104-3. Reprinted with permission from Radio-Electronics, 6/86, p. 60. © Copyright Gernsback Publications, Inc., 1986.

Fig. 104-4. Harris, Analog Product Data Book, 1988, p. 2 – 106.

Fig. 104-5. Reprinted with permission from Radio-Electronics, 6/86, p. 58. © Copyright Gernsback Publications, Inc., 1986.

Index

Numbers preceded by a "I," "II," "III," or "IV" are from *Encyclopedia of Electronic Circuits* Vol. I, II, III, or IV, respectively.

B

C

D

G

H

I

M

N

O

Q

R

S

T

U

V

W

X

Y

Z

Other Bestsellers of Related Interest

ENCYCLOPEDIA OF ELECTRONIC CIRCUITS Vol. 1—Rudolf F. Graf

". . . schematics that encompass virtually the entire spectrum of electronics technology . . . This is a well worthwhile book to have handy." ***—Modern Electronics***

Discover hundreds of the most versatile electronic and integrated circuit designs, all available at at the turn of a page. You'll find circuit diagrams and schematics for a wide variety of practical applications. Many entries also include clear, concise explanations of the circuit configurations and functions. 768 pages, 1,762 illustrations. Book No. 1938, $29.95 paperback, $60.00 hardcover

THE ILLUSTRATED DICTIONARY OF ELECTRONICS—5th Edition

—Rufus P. Turner and Stan Gibilisco

This completely revised and updated edition defines more than 27,000 practical electronics terms, acronyms, and abbreviations. Find up-to-date information on basic electronics, computers, mathematics, electricity, communications, and state-of-the-art applications—all discussed in a nontechnical style. The author also includes 360 new definitions and 125 illustrations and diagrams. 736 pages, 650 illustrations. Book No. 3345, $26.95 paperback, $39.95 hardcover

ELECTRONIC CONVERSIONS: Symbols and Formulas—2nd Edition

—Rufus P. Turner and Stan Gibilisco

This revised and updated edition supplies all the formulas, symbols, tables, and conversion factors commonly used in electronics. Exceptionally easy to use, the material is organized by subject matter. Its format is ideal and you can save time by directly accessing specific information. Topics cover only the most-needed facts about the most often used conversions, symbols, formulas, and tables. 280 pages, 94 illustrations. Book No. 2865, $14.95 paperback only

ELECTRONIC DATABOOK—4th Edition

—Rudolf F. Graf

If it's electronic, it's here—current, detailed, and comprehensive! Use this book to broaden your electronics information base. Revised and expanded to include all up-to-date information, this fourth edition makes any electronic job easier and less time-consuming. You'll find information that will aid in the design of local area networks, computer interfacing structure, and more! 528 pages, 131 illustrations. Book No. 2958, $24.95 paperback only

BUILD YOUR OWN TEST EQUIPMENT

—Homer L. Davidson

Build more than 30 common electronic testing devices, ranging from simple continuity and polarity testers to signal injectors and power supplies. Also learn how test instruments work, how they are used, and how to save money. Each project includes a complete parts list with exact part numbers. 300 pages, 324 illustrations. Book No. 3475, $17.95 paperback, $27.95 hardcover

ELECTRONIC SIGNALS AND SYSTEMS: Television, Stereo, Satellite TV, and Automotive—Stan Prentiss

Study signal analysis as it applies to the operation and signal-generating capabilities of today's most advanced electronic devices with this handbook. It explains the composition and use of a wide variety of test instruments, transmission media, satellite systems, stereo broadcast and reception facilities, antennas, television equipment, and even automotive electrical systems. You'll find coverage of C- and Ku-band video, satellite master TV systems, high-definition television, C-QUAM® AM stereo transmission and reception, and more. 328 pages, 186 illustrations. Book No. 3557, $19.95 paperback, $29.95 hardcover

MASTERING ELECTRONICS MATH —2nd Edition—R. Jesse Phagan

A self-paced text for hobbyists and a practical toolbox reference for technicians, this book guides you through the practical calculations needed to design and troubleshoot circuits and electronics components. Clear explanations and sample problems illustrate each concept, including how each is used in common electronics applications. If you want to gain a strong understanding of electronics math and stay on top of your profession, this book will be a valuable tool for you! 344 pages, 270 illustrations. Book No. 3589, $17.95 paperback, $27.95 hardcover

BOB GROSSBLATT'S GUIDE TO CREATIVE CIRCUIT DESIGN—Robert Grossblatt

Robert Grossblatt, *Radio Electronics'* popular columnist brings his unique circuit design philosophy and style to this hands-on guide. Emphasizing the importance of scientific method over technical knowledge, it walks you through the circuit design process—from brainwork to paperwork to boardwork—and suggests ways for making your bench time as efficient as possible. 248 pages, 129 illustrations. Book No. 3610, $17.95 paperback, $28.95 hardcover

SECRETS OF RF CIRCUIT DESIGN
—Joseph J. Carr

This book explains in clear, nontechnical language what RF is, how it works, and how it differs from other electromagnetic frequencies. You'll learn the basics of receiver operation, the proper use and repair components in RF circuits, and principles of radio signal propagation from low frequencies to microwave. You'll enjoy experiments that explore such problems as electromagnetic interface. 416 pages, 411 illustrations. Book No. 3710, $19.95 paperback, $32.95 hardcover

INTERNATIONAL ENCYCLOPEDIA OF INTEGRATED CIRCUITS—2nd Edition
—Stan Gibilisco

The most thorough coverage of foreign and domestic integrated circuits is available today in this giant resource. Seven separate sections detail thousands of ICs and their applications, including all relevant information, charts, and tables. This second edition of a unique all-in-one reference tells what each IC is, what it does, how it does it, and what its relationship is to other ICs and their applications. 1,168 pages, 4,605 illustrations. Book No. 3802, $84.95 hardcover only

ELECTRONIC POWER CONTROL: Circuits, Devices & Techniques—Irving M. Gottlieb

This guide focuses on the specific digital circuits used in electronic power applications. It presents state-of-the-art approaches to analysis, troubleshooting, and implementation of new solid-state devices. Gottlieb shows you how to adapt various power-control techniques to your individual needs. He uses descriptive analysis and real-world applications wherever possible, employing mathematical theory only when relevant. 272 pages, 197 illustrations. Book No. 3837, $17.95 paperback, $27.95 hardcover

ENCYCLOPEDIA OF ELECTRONICS —2nd Edition—Stan Gibilisco and
Neil Sclater, Co-Editors-in-Chief
Praise for the first edition:
". . . a fine one-volume source of detailed information for the whole breadth of electronics."

—Modern Electronics

The second edition brings you more than 950 pages of listings and cover virtually every electronics concept and component imaginable. From basic electronics terms to state-of-the-art applications, this is the most complete and comprehensive electronics reference available! 976 pages, 1,400 illustrations. Book No. 3389, $69.50 hardcover only

MASTER HANDBOOK OF ELECTRONIC TABLES AND FORMULAS—5th Edition
—Martin Clifford
". . . a source of quick, accurate, and easy-to-use solutions to electronics problems . . . to be used as a reference book for hobbyists as well as professionals."

Electronics for You, on a previous edition

It's a completely revised and updated edition of the classic reference used by thousands of hobbyists, technicians, and engineers. Reflecting the latest developments in the field, you'll never again have to stop and make pencil and paper calculations, hunt up your calculator for simple conversions, or search through reference books trying to find a specific formula. Everything you need is right here—logically organized and fully indexed. Martin Clifford includes the most current information on everything from resistance formulas, sine waves, capacitance, impedance vectors, decibels, and more. 544 pages, 490 illustrations. Book No. 3739, $22.95 paperback, $39.95 hardcover

Prices Subject to Change Without Notice.

Look for These and Other TAB Books at Your Local Bookstore

To Order Call Toll Free 1-800-822-8158

(in PA, AK, and Canada call 717-794-2191)

or write to TAB Books, Blue Ridge Summit, PA 17294-0840.

Title	Product No.	Quantity	Price

☐ Check or money order made payable to TAB Books

Charge my ☐ VISA ☐ MasterCard ☐ American Express

Acct. No. ____________ Exp. ________

Signature: ____________

Name: ____________

Address: ____________

City: ____________

State: ____________ Zip: ____________

Subtotal $ ________

Postage and Handling
($3.00 in U.S., $5.00 outside U.S.) $ ________

Add applicable state and local
sales tax $ ________

TOTAL $ ________

TAB Books catalog free with purchase; otherwise send $1.00 in check or money order and receive $1.00 credit on your next purchase.

Orders outside U.S. must pay with international money in U.S. dollars

TAB Guarantee: If for any reason you are not satisfied with the book(s) you order, simply return it (them) within 15 days and receive a full refund.

BC

WITH CUMULATIVE INDEX

Rudolf F. Graf
&
William Sheets

Encyclopedia of ELECTRONIC CIRCUITS

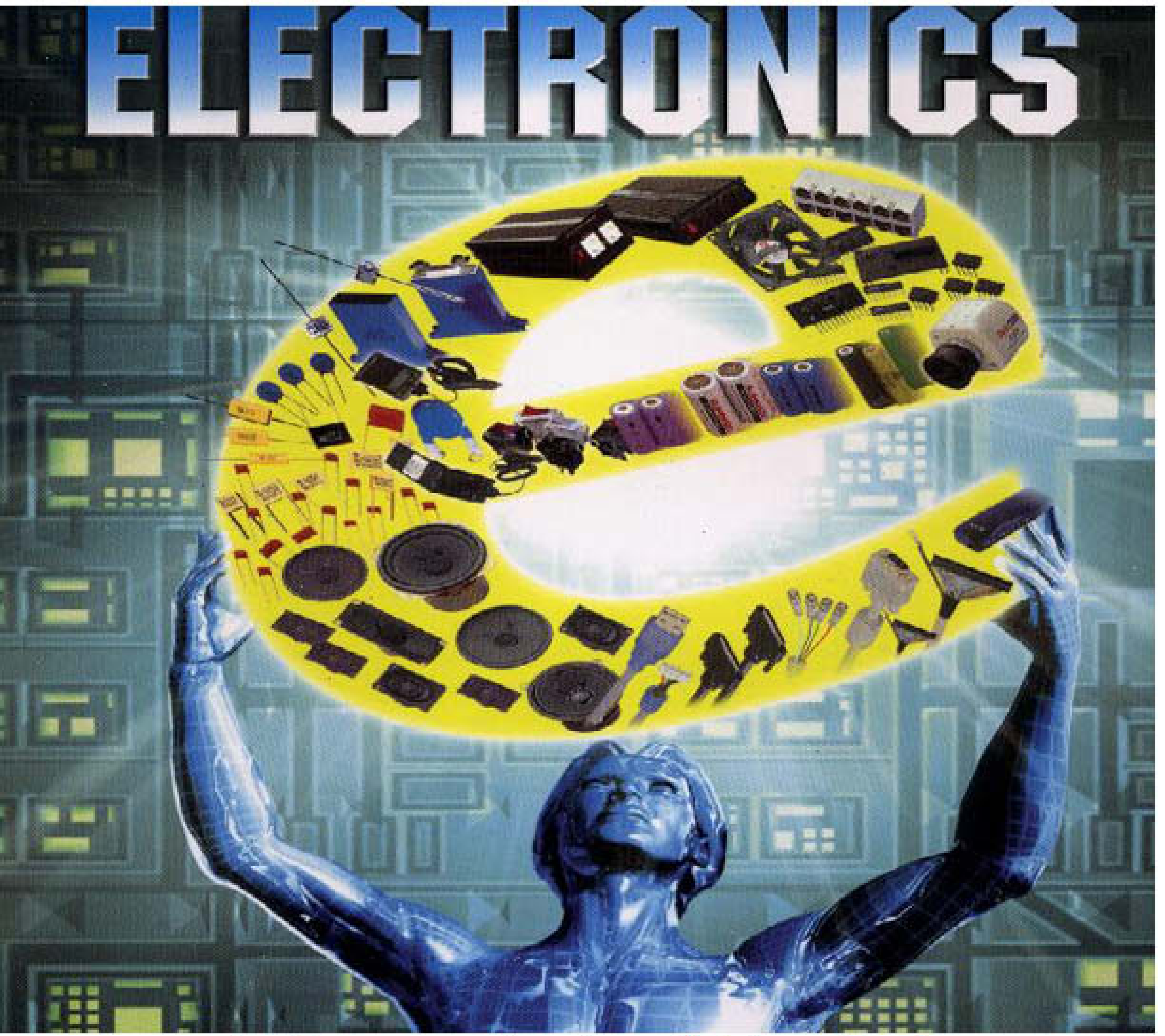
Volume 5

Encyclopedia of

ELECTRONIC CIRCUITS

Volume 5

Patent notice


ELECTRONICS

Encyclopedia of

ELECTRONIC CIRCUITS

Volume 5

Rudolf F. Graf
&
William Sheets

TAB Books
Division of McGraw-Hill, Inc.
New York San Francisco Washington, D.C. Auckland Bogotá
Caracas Lisbon London Madrid Mexico City Milan
Montreal New Delhi San Juan Singapore
Sydney Tokyo Toronto

Published by TAB Books, a division of McGraw-Hill, Inc.

pbk 5 6 7 8 9 10 11 12 13 / 9 9 8 7 6 5

Library of Congress Cataloging-in-Publication Data
(Revised for vol. 5)

Graf, Rudolf F.
The encyclopedia of electronics circuits.

Authors for v. 5– : Rudolf F. Graf & William Sheets.
Includes bibliographical references and indexes.
1. Electronic circuits—Encyclopedias. I. Sheets, William. II. Title.
TK7867G66 1985 621.3815 84-26772
ISBN 0-8306-0938-5 (v. 1)
ISBN 0-8306-1938-0 (pbk. : v. 1)
ISBN 0-8306-3138-0 (pbk. : v 2)
ISBN 0-8306-3138-0 (v. 2)
ISBN 0-8306-3348-0 (pbk. : v. 3)
ISBN 0-8306-7348-2 (v. 3)
ISBN 0-8306-3895-4 (pbk. : v. 4)
ISBN 0-8306-3896-2 (v. 4)
ISBN 0-07-011077-8 (pbk. : v. 5)
ISBN 0-07-011076-X (v. 5)

Acquisitions Editor: Roland S, Phelps
Editorial team: Andrew Yoder, Book Editor
Joanne Slike, Executive Editor
Production Team: Katherine G. Brown, Director
Jan Fisher, Coding
Lisa M. Mellott, Coding
Rose McFarland, Layout
Linda L. King, Proofreading
Nancy K. Mickley Proofreading
Joann Woy, Indexer
Design team: Jaclyn J. Boone, Designer
Brian Allison, Associate Designer
Cover design: Stickles Associates, Bath, Pa.

EL1
0110778

Contents

Introduction

The *Encyclopedia of Electronic Circuits, Volume V* adds approximately 1000 new circuits to the treasury of carefully chosen circuits that cover nearly every phase of today's electronic technology. These five volumes contain a wealth of new ideas and up-to-date circuits garnered from prestigious industry sources. Also included are some of the authors' original designs.

Each circuit is accompanied by a brief explanation of how it works, unless the circuit's operation is either obvious or too complex to describe in a few words. In the latter case, the reader should consult the original source listed in the back of the book. The index includes all entries from Volumes I to V. This provides instant access to about 5000 circuits, which make up the most extensive collection of carefully categorized modern circuits available anywhere.

Once again, the authors wish to extend their thanks to Ms. Loretta Gonsalves, whose virtuoso performance at the word processor contributed so much to the successful completion of the manuscript for this work. We look forward to the pleasure of working with her on Volume VI, which is now under development.

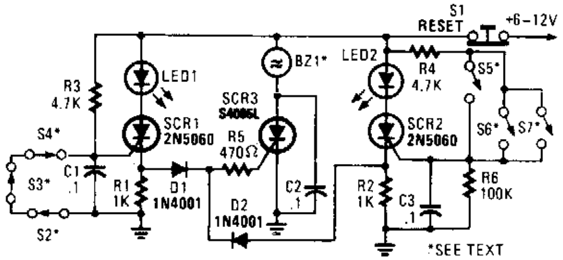
Rudolf F. Graf and William Sheets

1

Alarm and Security Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

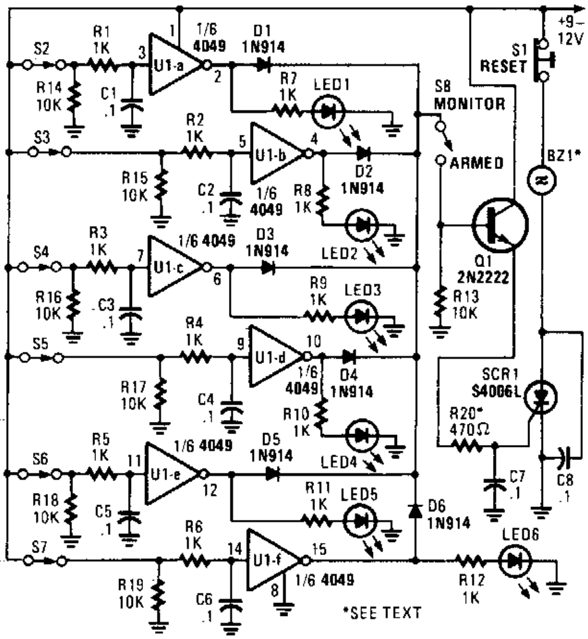
HIGH-POWER ALARM DRIVER



In this circuit, a low-powered SCR is used to trigger a higher powered SCR. When a switch is opening (S2, S3, S4) or closing (S5, S6, S7), either SCR1 or SCR2 triggers. This triggers SCR3 via D1, D2, and R5. BZ1 is a high-powered alarm of the noninterrupting type.

POPULAR ELECTRONICS ***FIG. 1-1***

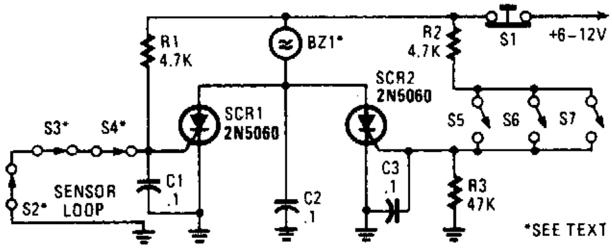
MULTI-LOOP PARALLEL ALARM



POPULAR ELECTRONICS ***FIG. 1-2***

This alarm has status LEDs connected across each inverter output to indicate the status of its associated sensor. S8 is used to monitor the switches via the LEDs, or to trigger an alarm via Q1 and SCR1. BZ1 should be a suitable alarm of the noninterrupting type.

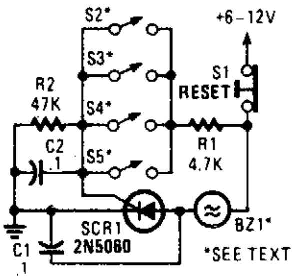
SERIES/PARALLEL LOOP ALARM



POPULAR ELECTRONICS *FIG. 1-3*

Two SCRs are used with two sensor loops. One loop uses series switches, the other loop parallel switches. When a switch actuation occurs, the SCR triggers. The alarm should be a noninterrupting type.

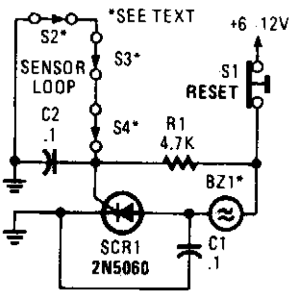
PARALLEL LOOP ALARM



POPULAR ELECTRONICS *FIG. 1-4*

Four parallel switches are used to monitor four positions. When a closure occurs on any switch, SCR1 triggers, which sounds the alarm. The alarm should be of the noninterrupting type.

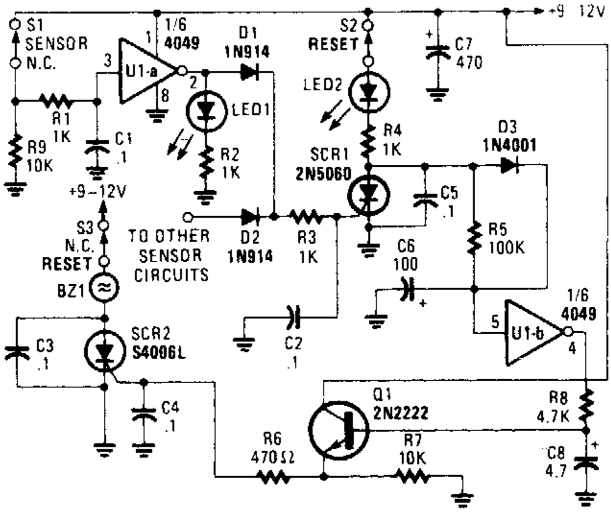
CLOSED-LOOP ALARM



POPULAR ELECTRONICS *FIG. 1-5*

A string of three series-connected, normally closed switches are connected across the gate of an SCR. When one opens, the SCR triggers via R1, sounding an alarm. The alarm should be of the noninterrupting type.

DELAYED ALARM



POPULAR ELECTRONICS

FIG. 1-6

The alarm/sensor circuit shown is built around two SCRs, a transistor, a 4049 hex inverter, and a few support components, all of which combine to form a closed-loop detection circuit with a delay feature. The delay feature allows you to enter a protected area and deactivate the circuit before the sounder goes off.

Assuming that the protected area has not been breached (i.e., S1 is in its normally-closed position), when power is first applied to the circuit, a positive voltage is applied to the input of U1-a through S1 and R1, causing its output to go low. That low is applied to the gate of SCR1, causing it to remain off. At the same time, C6 rapidly charges toward the +*V* supply rail through S2, LED2, R4, and D3. The charge on C6 pulls pin 5 of U1-b high, causing its output at pin 4 to be low. That low is applied to the base of Q1, keeping it off. Because no trigger voltage is applied to the gate of SCR2 (via Q1), the SCR remains off and BZ1 does not sound.

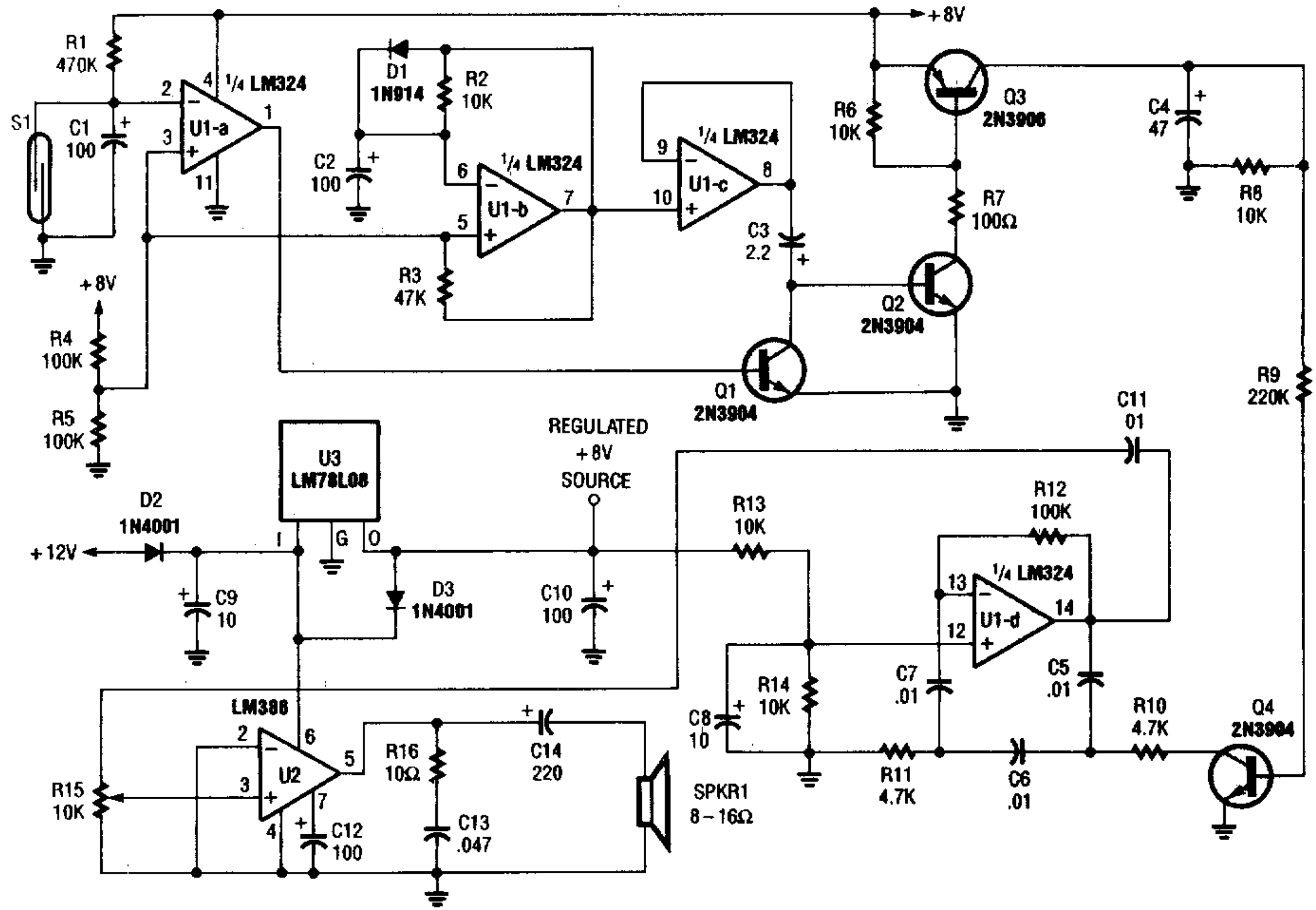
But should S1 open, the input of U1-a is pulled low via R9, forcing the output of U1-a high, lighting LED1. That high is also applied to the gate of SCR1 through D1 and R3, causing SCR1 to turn on. With SCR1 conducting, the charge on C6 decays, the input of U1-b at pin 5 is pulled low, forcing its output high, slowing charging C8 through R8 to a voltage slightly less than the positive supply rail.

Transistor Q1 remains off until C8 has charged to a level sufficient to bias Q1 on, allowing sufficient time to enter the protected area and disable the alarm before it sounds. Once C8 has developed a sufficient charge, Q1 turns on and supplies gate current to SCR2 through R6, causing the SCR to turn on and activate BZ1. If the circuit is reset before the delay has timed out, no alarm will sound.

The delay time can be lengthened by increasing the value of either or both C6 and R5; decreasing the value of either or both of those components will shorten the delay time.

All of the switches used in the circuit are of the normally-closed (NC) variety. Switch S1 can be any type of NC security switch. Switch S2 can be either a pushbutton or toggle switch. Because S3 is used to disable the sounder (BZ1) only, anything from a key-operated security switch to a hidden toggle switch can be used.

DOOR MINDER

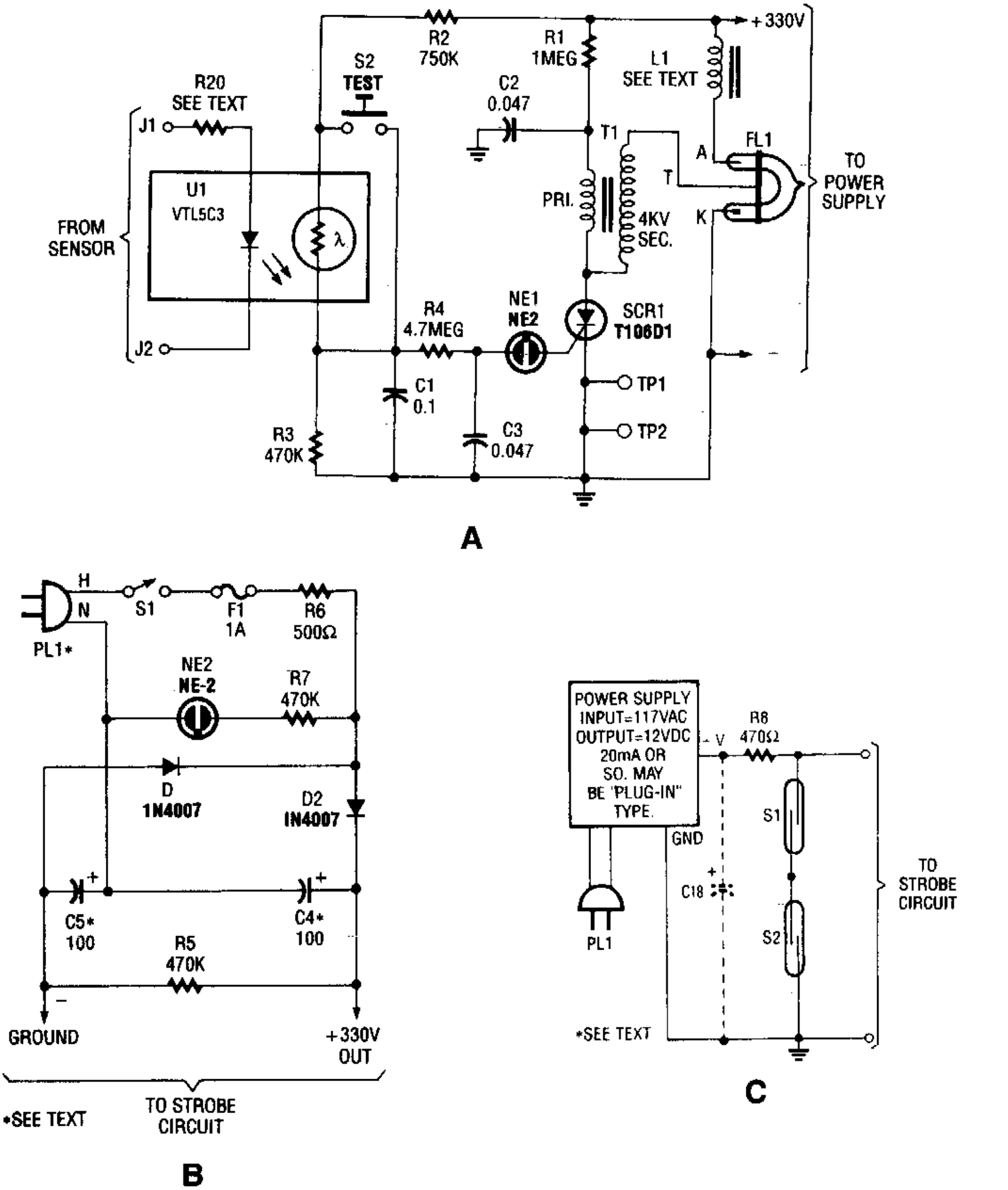


POPULAR ELECTRONICS

FIG. 1-7

This circuit monitors a door to determine if it has been left open. After 24 seconds, the alarm sounds. S1 is a magnetic sensor. The alarm is an electronic chime sound that is struck once per second.

STROBE ALERT SYSTEM



1992 PE HOBBYIST HANDBOOK ***FIG. 1-8***

The circuit is activated by an LED/photoresistor isolator (U1), which is a combination of a light-dependent resistor (LDR) and an LED in a single package. That device was chosen because of its high isolation (2000 V) characteristic, which is necessary because the strobe part of the circuit is directly connected to the ac line.

STROBE ALERT SYSTEM (*Cont.*)

The voltage divider is formed by R2, U1's internal resistance, and R3. When U1's internal LED is off, U1's internal LDR has a very high resistance—on the order of 10 MΩ. The voltage applied to NE1 is considerably below its ignition voltage of approximately 90 Vdc.

The optoisolator's internal LED is activated by a dc signal supplying 20 mA. The external sensor(s) that supply the signal are connected to the strobe part of the circuit at J1 and J2.

When the internal LED lights, the LDR's resistance decreases to around 5 kΩ. Under that condition, about 125 Vdc is applied across C1, R4, and C2. The neon lamp periodically fires and extinguishes as capacitor C3 charges through R4, and discharges via NE1 and the SCR gate.

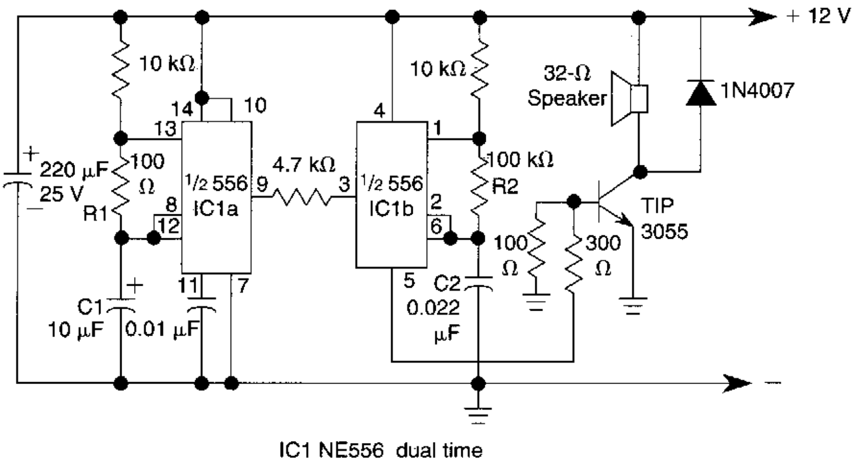
Resistor R4 restricts the current input to C3, and thereby controls the firing rate of NE1—about three times per second. The discharge through NE1 is applied to the gate of SCR1.

SCR1, a sensitive-gate unit, snaps on immediately when NE1 conducts, which completes the ground circuit for transformer T1 (a 4-kV trigger transformer). As SCR1 toggles on and off in time with the firing of NE1, capacitor C2 (connected in parallel with T1's primary) charges via R1, and then discharges very rapidly through T1's primary winding. A voltage pulse is applied to the trigger input of FL1, a Xenon flash lamp.

It is important to remember that the circuit is connected directly to the ac line. Resistor R6 is included to limit the amount of line current available to the circuit. The value of R6 can be decreased if you intend to modify the circuit for more flash power.

Warning: Even though the circuit is fuse-protected, it can still be dangerous if handled carelessly.

WARBLE ALARM

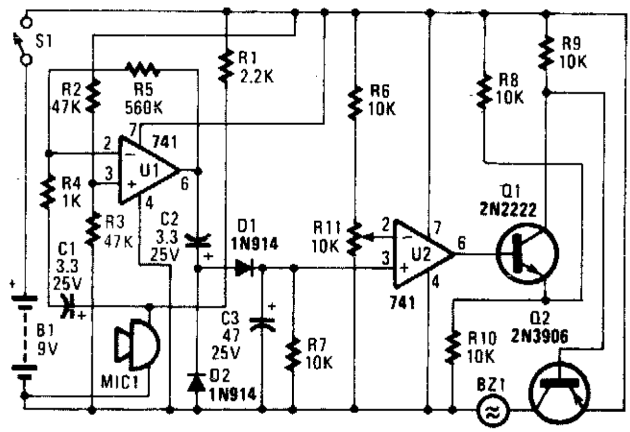


WILLIAM SHEETS

FIG. 1-9

This circuit uses a 556 to first generate a low frequency square wave, that is modulated to produce two alternate tones of about 400 and 500 Hz. Circuit generates warble alarm of European emergency vehicles. The frequencies of the oscillators are determined by the values of R1, C1 and R2, C2.

AUDIO ALARM



POPULAR ELECTRONICS **FIG. 1-10**

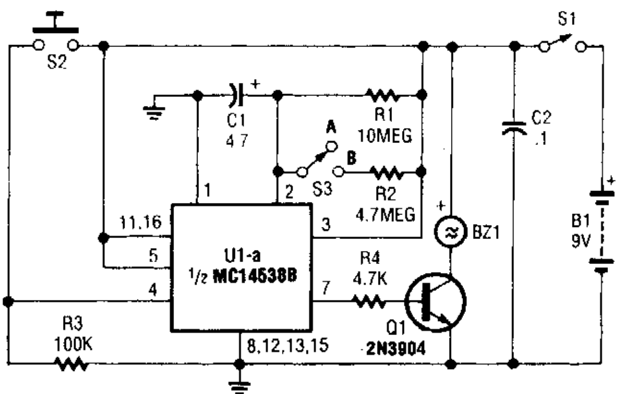
In the circuit, U1 amplifies the audio picked up by the condenser microphone. Resistor R1 limits current, while R2 and R3 center the output of the amplifier to ½*B*+ to allow a single-ended supply to be used. Diodes D1 and D2 rectify the output of U1, and C3 filters the resulting pulsing dc. Thus, a dc voltage that is proportional to the ambient sound level is produced.

That voltage is presented to the noninverting input of U2. The inverting input is provided with a reference voltage of between 0 and ½*B*+, which is set by R11.

As long as the noise level is low enough to keep the voltage at pin 3 lower than the voltage at pin 2, the output of U2 stays low (approximately 1 V). That is enough to bias Q1 partially on. A voltage divider, formed by R8/R10 and Q1 (when it's partially on), prevents Q2 from turning on.

When the noise level is high enough to bring the voltage at pin 3 higher than the voltage at pin 2, the output of U2 goes high. That turns Q1 fully on and drives Q2 into saturation. The piezo buzzer then sounds until the power is cut off.

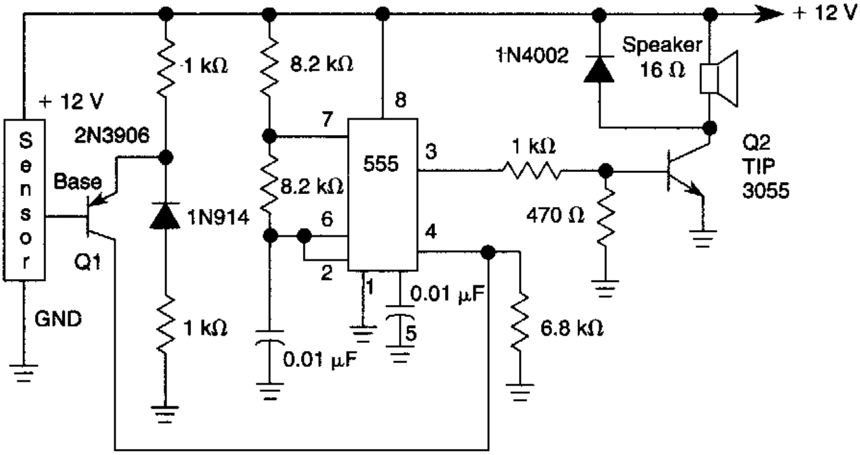
NO-DOZE ALARM



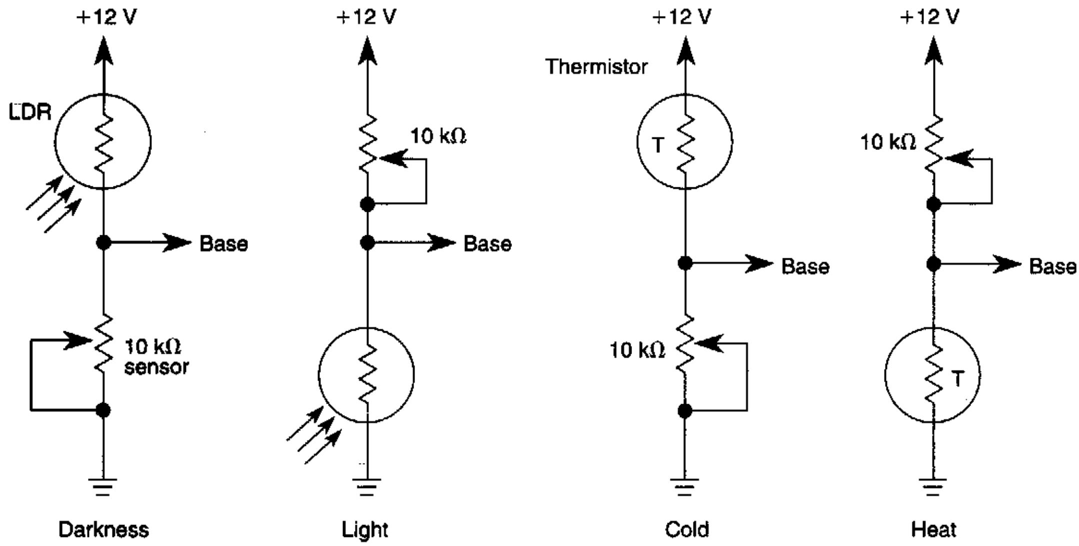
This circuit sends out a loud tone if the input switch (S2) is not retriggered at preset intervals. If you fall asleep and miss retriggering the circuit, it will sound until you press S2.

POPULAR ELECTRONICS **FIG. 1-11**

HEAT- OR LIGHT-ACTIVATED ALARM



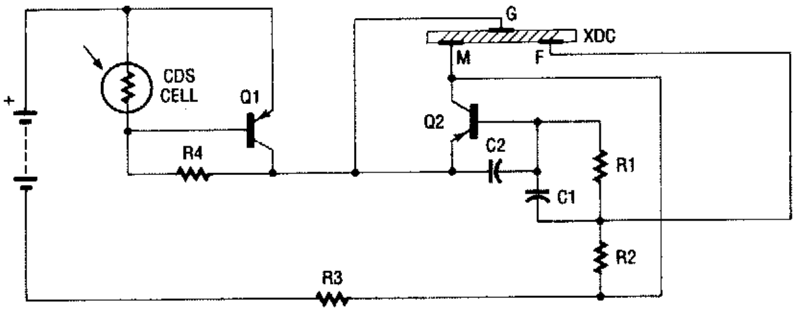
SENSOR CIRCUITS



WILLIAM SHEETS ***FIG. 1-12***

The tone generated by a 555 oscillator can be turned on (activated) by heat or light. That causes Q1 to conduct transistor W2 (TIP 3055). Q2 (TIP 3055) acts as an audio amplifier and speaker driver.

PIEZOELECTRIC ALARM



C1,C2	.047 Disc Capacitor
Q1,Q2	2N3906 Transistor
R1	6.8K Resistor
R2	33K Resistor
R3	200 ohm Resistor
R4	75K Resistor
XDC	Piezoelectric transducer

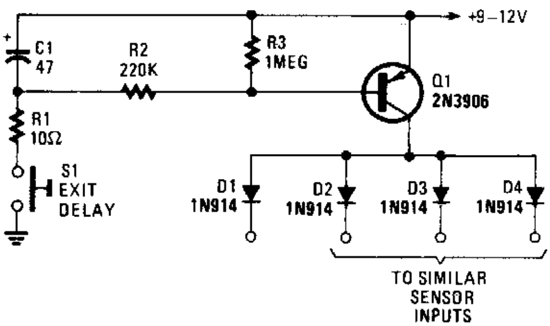
1991 PE HOBBYIST HANDBOOK ***FIG. 1-13***

The alarm uses a fixed-frequency piezoelectric buzzer in conjunction with the cadmium-sulfide (CDS) cell and the two-transistor circuit to provide a unique effect. Whenever light reaches the CDS photo-electric cell, the alarm is silent. But when no light strikes the cell, transistor Q1 turns on, and the circuit emits a high-pitched tone.

The alarm consists of a piezoelectric disk that oscillates at the fixed frequency of 3.137 kHz, created by transistor Q2, capacitor C1 and C2, and resistors R1 through R3. Transistor Q1 is used as a switch. It is forward-biased "on" by R4; however, the CDS cell turns Q1 "off" when the light is striking it.

A CDS photo cell is made from cadmium sulfide, a semiconductor material that changes resistance when the light strikes it. The greater the amount of light, the lower the resistance. The low resistance conducts positive voltage to the base of pnp transistor Q1, keeping it turned "off" when the light shines on the CDS cell. As soon as the light is removed, the CDS cell provides a resistance of over 100 kΩ. That causes Q1 to turn "on," allowing a positive voltage to reach the emitter lead of Q2, which then begins to oscillate. That then causes the piezoelectric element (transducer) to produce a loud signal.

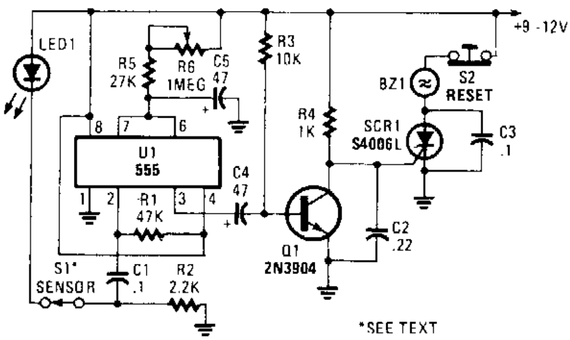
EXIT DELAY FOR BURGLAR ALARMS



Depressing S1 charges C1 to the supply voltage. This biases Q1 on via bias resistors R2 and R3. A voltage is available for the duration of the delay period, to hold off the alarm circuit. C1 can be increased or decreased in value to alter the delay times.

POPULAR ELECTRONICS ***FIG. 1-14***

555-BASED ALARM



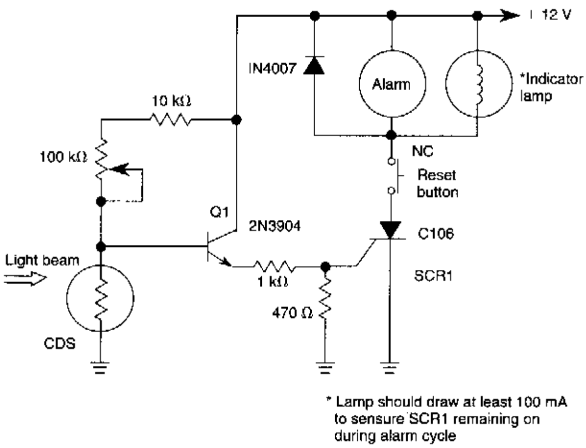
POPULAR ELECTRONICS *FIG. 1-15*

The alarm circuit has a single 555 oscillator/timer (U1) performing double duty; serving both in the alarm-trigger circuit and the entry-delay circuit. In this application, the trigger input of U1 at pin 2 is held high via R1. A normally-closed sensor switch, S1, supplies a positive voltage to the junction of R2 and C1, and lights LED1. With both ends of C1 tied high, there is no charge on C1. But when S1 opens, C1 (initially acting as a short) momentarily pulls pin 2 of U1 low, triggering the timed delay circle.

At the beginning of the timing cycle, U1 produces a positive voltage at pin 3, which charges C4 to near the positive voltage at pin 3, which charges C4 to near the positive supply voltage. Transistor Q1 is heavily biased on by R3, keeping its collector at near ground level. With Q1 on, SCR1's gate is clamped to ground, holding it off. When the delay circuit times out, pin 3 of U1 goes low and ties the positive end of C4 to ground. That turns Q1 off.

When Q1 turns off, the voltage at the gate of SCR goes positive, turning on the SCR and sounding the alarm. The delay time is adjustable from just a few seconds (R6 set to its minimum resistance) to about one minute (R6 adjusted to its maximum resistance).

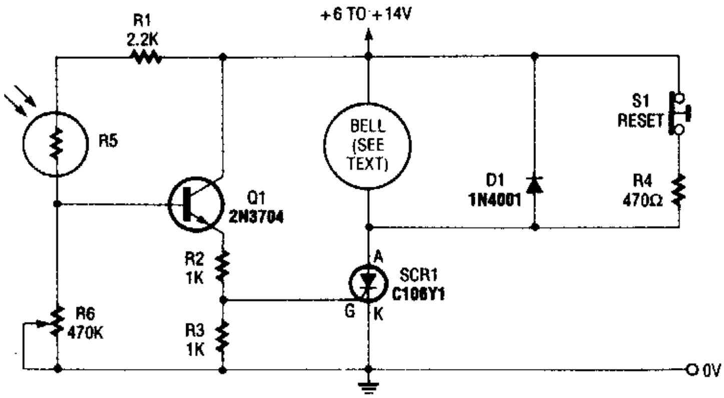
LIGHT-BEAM ALARM FOR INTRUSION DETECTION



When the light beam that falls in the CDS photocell is interrupted, transistor (EN3904) conducts thereby triggering SCR1 (C106) and activating alarm bell. S1 resets the SCR. The alarm bell should be a self-interrupting electromechanical type.

WILLIAM SHEETS *FIG. 1-16*

LIGHT-ACTIVATED ALARM WITH LATCH

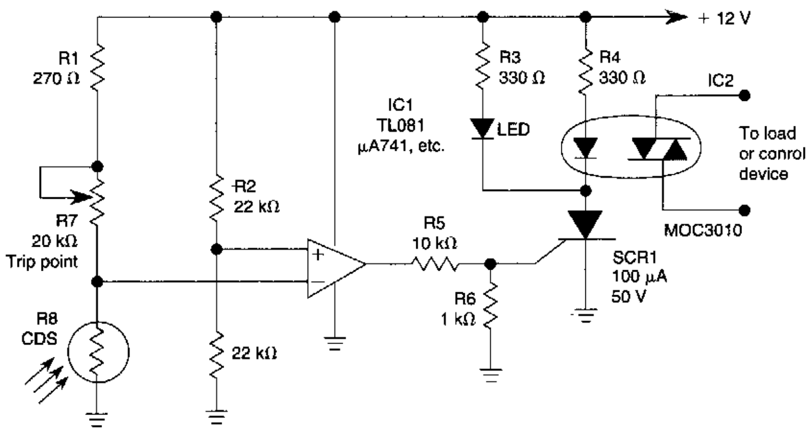


ELECTRONICS NOW

FIG. 1-17

In this circuit, light causes R5 to conduct forward-biasing Q1. R6 sets sensitivity. SCR1 is triggered from the emitter voltage on LQ1, sounding the alarm bell. When S1 is depressed, SCR1 unlatches. Be sure that a self-interrupting alarm (electromechanical buzzer or bell) is used.

PRECISION LIGHT-ACTIVATED ALARM

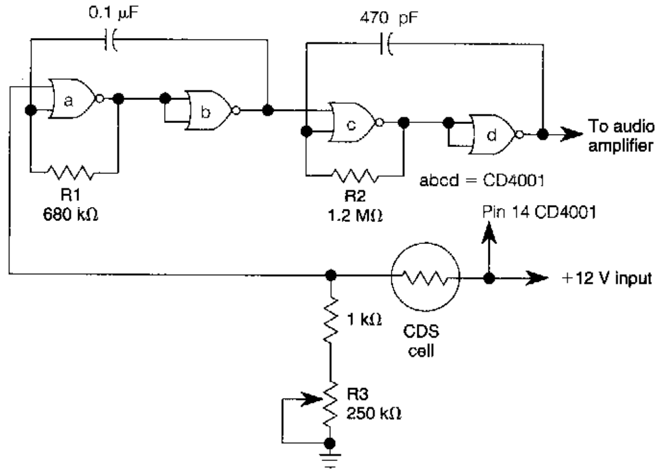


WILLIAM SHEETS

FIG. 1-18

The light-sensitive CDS cell R8 configured in a bridge circuit with IC1 as a comparator causes IC1's output to go high when light strikes the CDS cell R8, triggering SCR1. This lights LED1 and turns on opto isolator IC2, which switches the load.

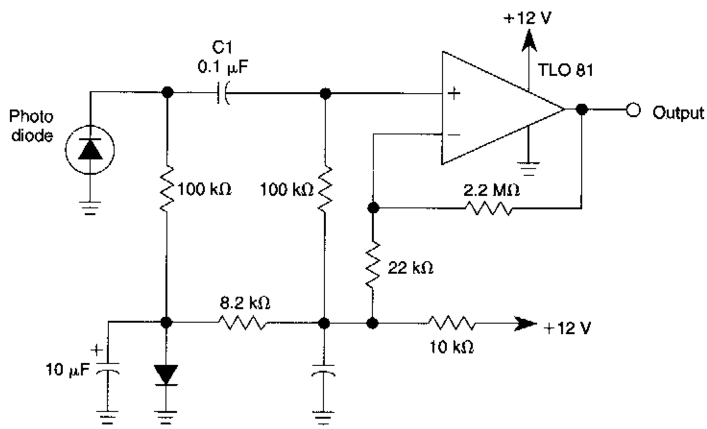
DARK-ACTIVATED ALARM WITH PULSED TONE OUTPUT



WILLIAM SHEETS **FIG. 1-19**

NOR gates a and b form a low-frequency oscillator that is activated when the CDS cell, under dark conditions, causes NOR gate a to see a logic zero at one input. This low-frequency (10 Hz) gates a high-frequency oscillator (c and d) to oscillate at around 1000 Hz. R1 can be varied to change the pulse rate and R2 to change the tone. R3 sets the trigger point.

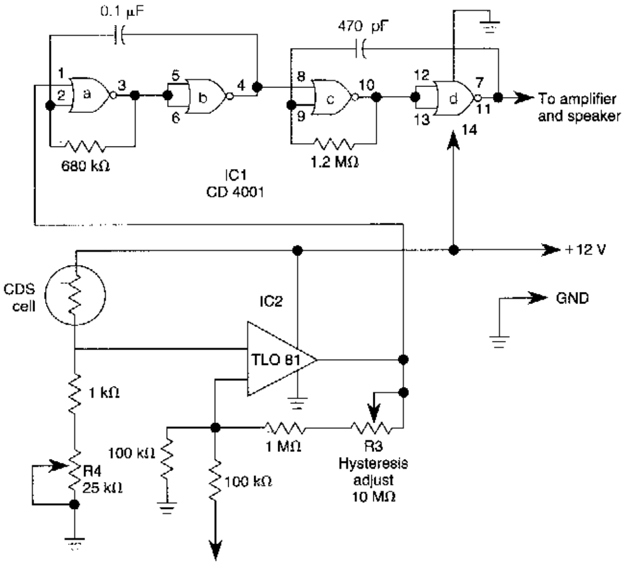
LIGHT-BEAM ALARM PREAMPLIFIER



WILLIAM SHEETS **FIG. 1-20**

This circuit can be used for light beams to 20 kHz. The gain of the operational amplifier is set for a 40-dB gain.

PRECISION LIGHT ALARM WITH HYSTERESIS

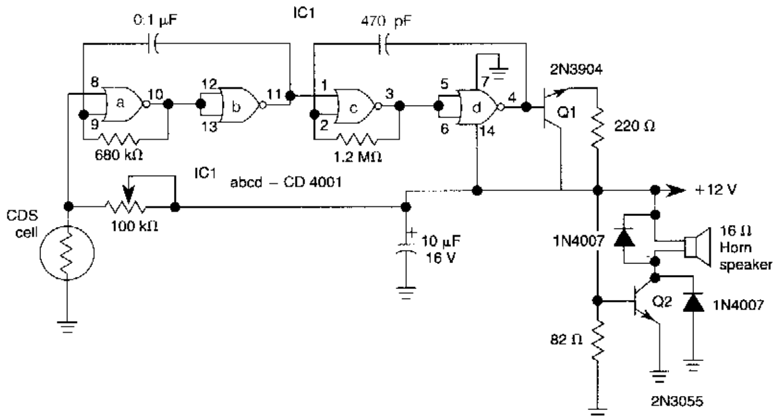


WILLIAM SHEETS

FIG. 1-21

The TL081 is used as a comparator in a Wheatstone bridge circuit. When the CDS cell resistance decreases due to exposure to light, the output from IC2 cause the low-frequency oscillator (a) and (b) to generate a 10-Hz square wave, gating the 1000 Hz oscillator (c) and (d) on and off. This signal drives an amplifier. R3 controls hysteresis, which reduces on-off triggering near the threshold set by R4.

HIGH-OUTPUT PULSED-TONE/LIGHT-ACTIVATED ALARM

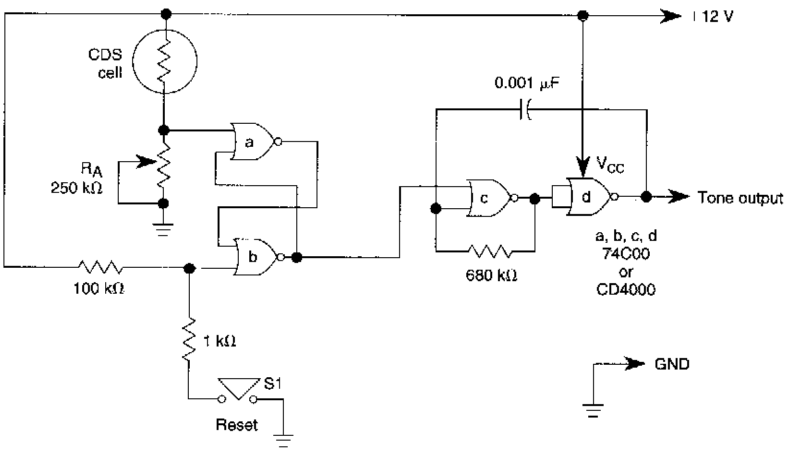


WILLIAM SHEETS

FIG. 1-22

This circuit can produce up to 1 W of audio power to drive a speaker or horn. When the CDS cell is struck by light, its resistance decreases thus activating NOR gate (a) thereby causing (a) and (b) to produce a low-frequency (10-Hz) square wave. This pulses the 1-kHz oscillator (c) and (d), causing it to generate a pulsed 1-kHz tone at a 10-Hz rate. Q1 and Q2 amplify this signal. Q2 (2N3055) drives the speaker.

SELF-LATCHING LIGHT ALARM WITH TONE OUTPUT

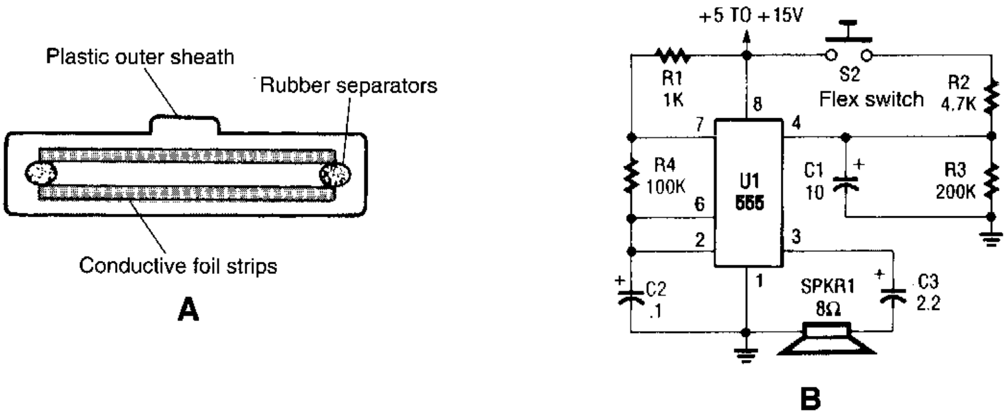


WILLIAM SHEETS

FIG. 1-23

A decrease in the resistance of the CDS cell when light strikes it activates latch a and b, enabling tone oscillator c and d which produces an output of about 1000 Hz. R_A sets the trip level. S1 resets the circuit.

ALARM SOUNDER FOR FLEX SWITCH

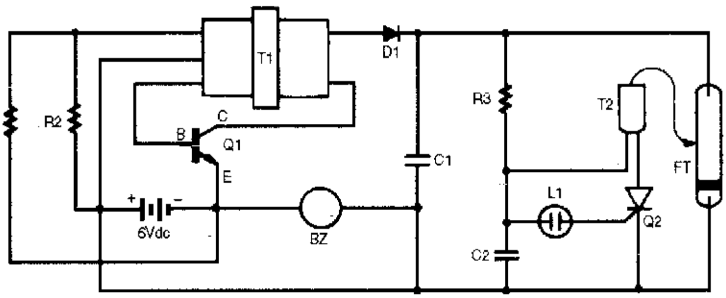


POPULAR ELECTRONICS

FIG. 1-24

This is a cross-sectional diagram of a flex switch. They can be used as pushbutton or even position sensors. This schematic diagram shows an oscillator, which is used as an alarm sounder, triggered by a flex switch.

BURGLAR CHASER

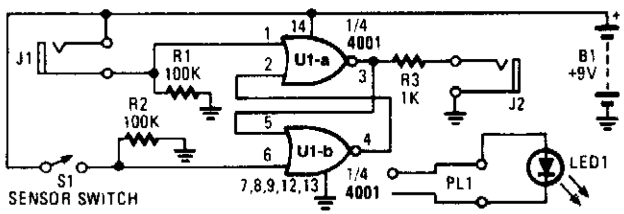


BZ .... Metal Horn Buzzer
C1 .... .5 μF 250 volts Capacitor
C2 .... .022 μF Green Cap (223 K5K)
D1 .... 1N4007 Diode
FT .... Micro Strobe Tube/Reflector
L1 .... Neon Lamp
Q1 .... C1740 SW Transistor
Q2 .... 106 SCR
R1 .... 200 ohm Resistor
R2 .... 820 ohm Resistor
R3 .... 10 meg Resistor
T1 .... Inverter Transformer
T2 .... 4 kV Trigger Coil

1991 PE HOBBYIST HANDBOOK ***FIG. 1-25***

The burglar chaser makes a great accessory for any alarm system. It creates brilliant flashes of white light and a loud, irritating sound from a metal horn buzzer. Transformer T1 is connected to Q1, R1, and R2 to form a blocking oscillator. This creates a 6-Vac signal on the primary of T1. Because of T1's large ratio of turns from primary to secondary, the 6-Vac signal is stepped up to a level of over 200 Vac, which is then rectified by D1. The resultant dc voltage is applied to storage capacitor C1 and the neon relaxation oscillator made up of R3, C2, and L1. Each time C2 charges up to a sufficient level, it ionizes L1, which causes SCR Q2 to fire. The firing SCR causes the charge on C2 to be applied to the trigger coil. The trigger coil converts the 200 V into the 4000-V pulse that is needed to fire micro xenon strobe tube/reflector FT. The cycle repeats itself after the strobe tube flashes.

SILENT ALARM



POPULAR ELECTRONICS ***FIG. 1-26***

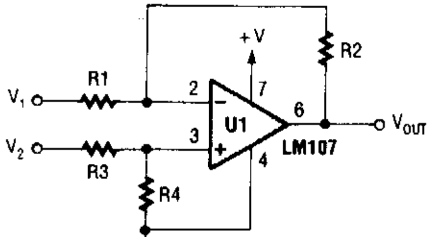
A sensor switch triggers a set-reset flip flop and lights an LED.

2

Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

DIFFERENCE AMPLIFIER



$$V_{OUT} = \frac{R3 + R2}{R1 + R4}\,\frac{R4}{R1}\,V_2 - \frac{R2}{R1}\,V_1$$

FOR R1 = R3 AND R2 = R4

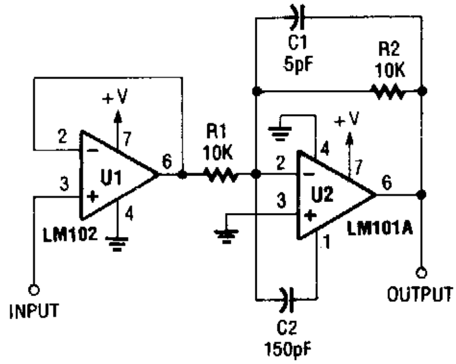
$$V_{OUT} = \frac{R2}{R1}\,(V_2 - V_1)$$

R1||R2=R3||R4

POPULAR ELECTRONICS ***FIG. 2-1***

By using two inputs as shown, a difference amplifier yielding the differential between U1 and U2, times a gain factor results.

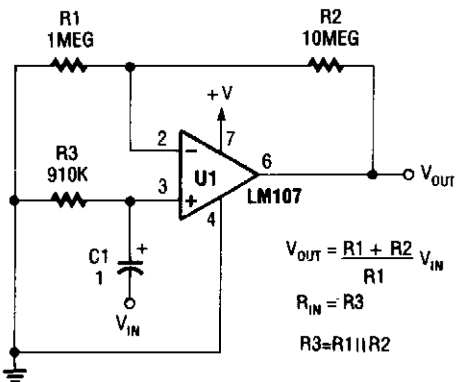
FAST-INVERTING AMPLIFIER WITH HIGH INPUT IMPEDANCE



POPULAR ELECTRONICS ***FIG. 2-2***

U1 is used as a voltage follower to feed inverter U2. Because U1 is in the voltage-follower configuration, it exhibits a high input impedance.

NONINVERTING ac AMPLIFIER



$$V_{OUT} = \frac{R1 + R2}{R1}\,V_{IN}$$

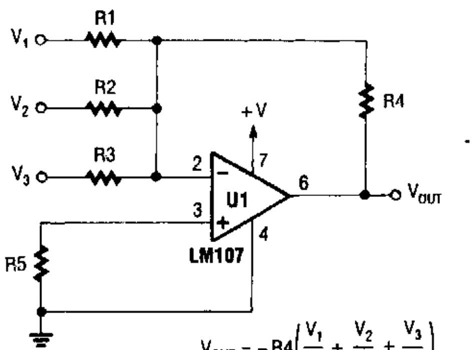
$R_{IN} = R3$

R3=R1||R2

POPULAR ELECTRONICS ***FIG. 2-3***

A general-purpose noninverting ac amplifier for audio of other low-frequency applications is shown. Design equations are in the figure. Almost any general-purpose op amp can be used for U1.

INVERTING SUMMING AMPLIFIER



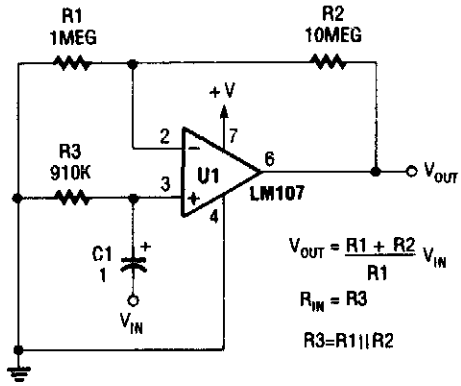
$$V_{OUT} = -R4\left(\frac{V_1}{R1} + \frac{V_2}{R2} + \frac{V_3}{R3}\right)$$

R5= R1||R2||R3||R4

POPULAR ELECTRONICS ***FIG. 2-4***

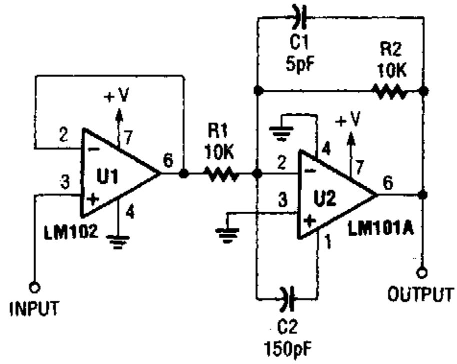
The output of U1 is the sum of V_1, V_2, and V_3, multiplied by R_1/R_4, R_2/R_4, and respectively. R1, R2, R3 are selected as required for individual gains. R4 affects gain of all these inputs.

NONINVERTING ac AMPLIFIER



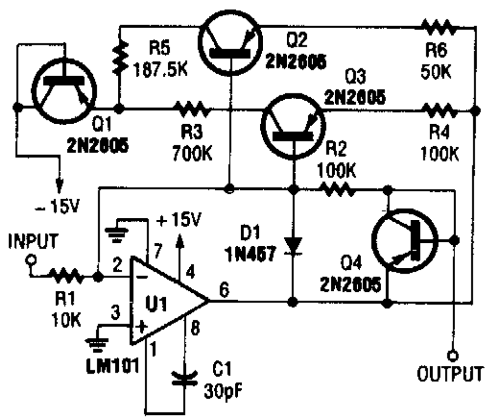
POPULAR ELECTRONICS *FIG. 2-5*

FAST HIGH-IMPEDANCE INPUT-INVERTING AMPLIFIER



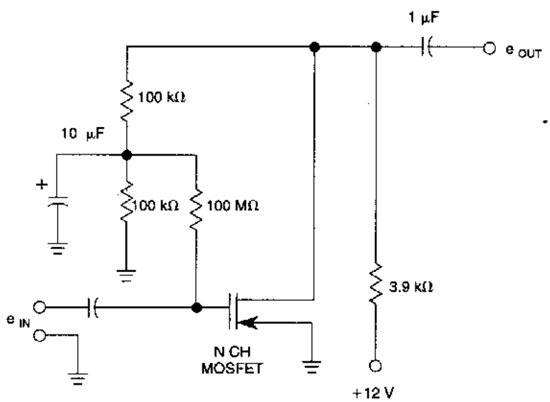
POPULAR ELECTRONICS *FIG. 2-6*

NONLINEAR OPERATIONAL AMPLIFIER WITH TEMPERATURE COMPENSATED-BREAKPOINT



POPULAR ELECTRONICS *FIG. 2-7*

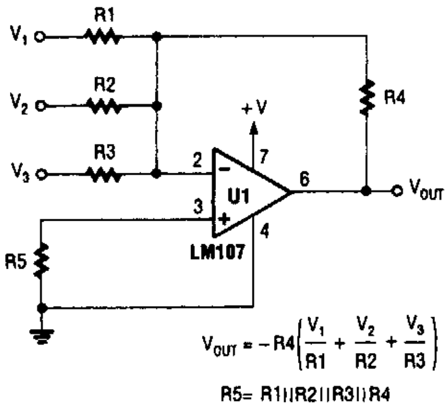
MOSFET HIGH-IMPEDANCE BIASING METHOD



WILLIAM SHEETS *FIG. 2-8*

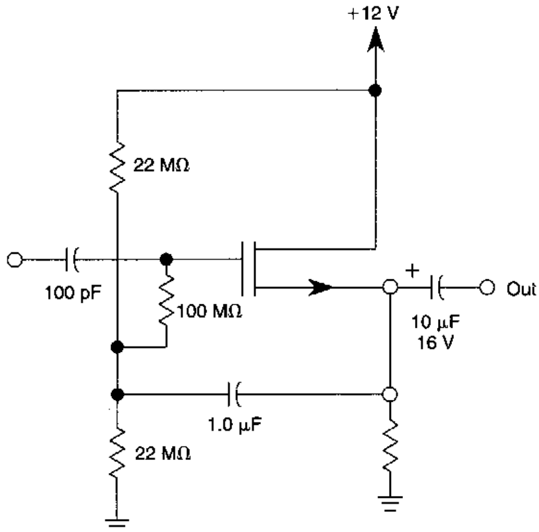
High-impedance biasing method for an N-channel MOSFET to form a linear-inverting amplifier.

INVERTING SUMMING AMPLIFIER



POPULAR ELECTRONICS ***FIG. 2-9***

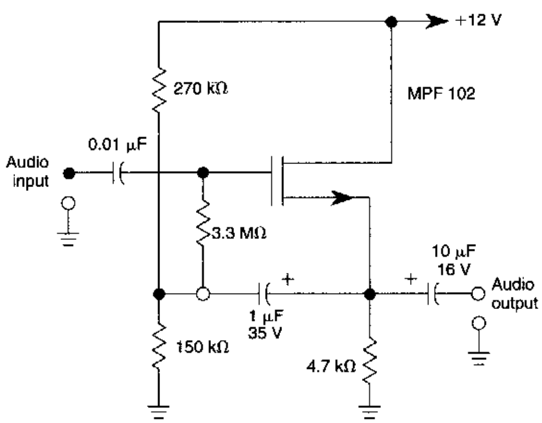
BOOTSTRAPPED SOURCE FOLLOWER



WILLIAM SHEETS ***FIG. 2-10***

This bootstrapped source follower uses an N-channel MOSFET. It has a high input impedance.

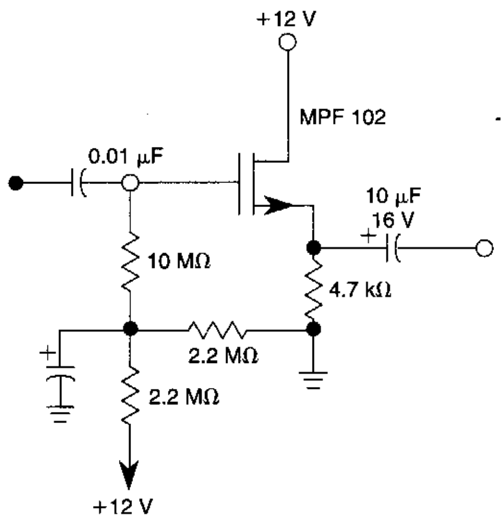
30-MΩ JFET SOURCE FOLLOWER



WILLIAM SHEETS ***FIG. 2-11***

This JFET source-follower uses an MPF102 with offset biasing. It has an input impedance of >30 MΩ.

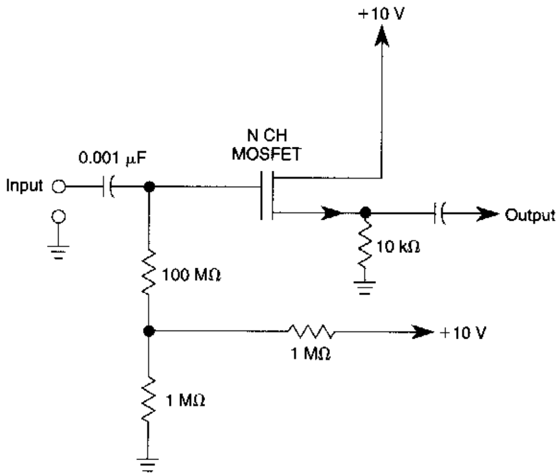
JFET SOURCE FOLLOWER



WILLIAM SHEETS ***FIG. 2-12***

The circuit uses positive gate bias to improve the operating point for better dynamic range.

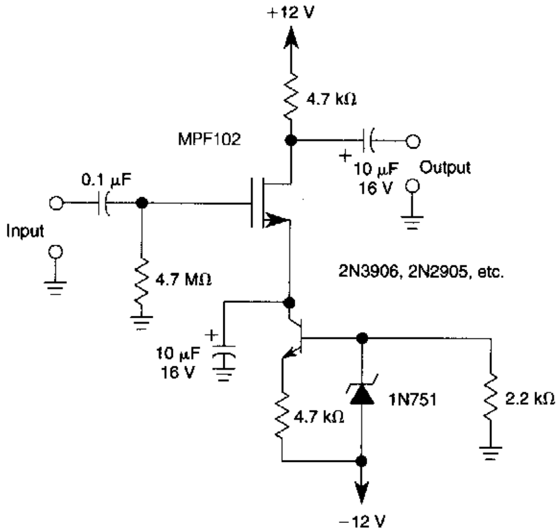
UNITY-GAIN NONINVERTING AMPLIFIER



WILLIAM SHEETS ***FIG. 2-13***

Biasing methods for an N-channel MOSFET to form a unity-gain noninverting amplifier or source-follower.

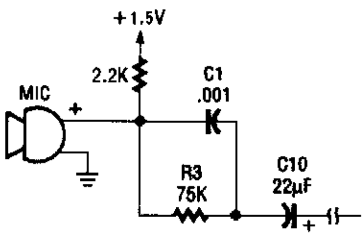
JFET AMP WITH CURRENT SOURCE BIASING



WILLIAM SHEETS ***FIG. 2-14***

A current source (MPF102) in the source lead of bipolar transistor 2N3906 permits accurate control of drain current.

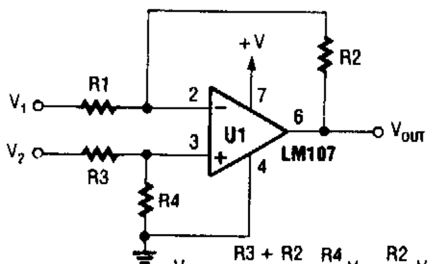
ELECTRET MIKE PREAMP



ELECTRONICS NOW ***FIG. 2-15***

This circuit is suitable for using an electret microphone for many applications. A 1.5-V battery is used. C1 and R3 provide treble boost/bass cut; they can be eliminated, if desired.

DIFFERENCE AMPLIFIER



$$V_{OUT} = \frac{R3 + R2}{R1 + R4}\,\frac{R4}{R1}\,V_2 - \frac{R2}{R1}\,V_1$$

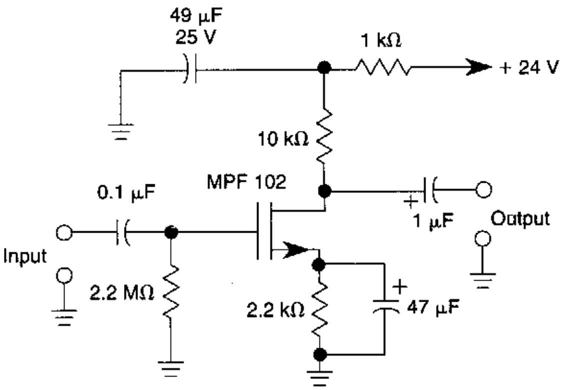
FOR R1 = R3 AND R2 = R4

$$V_{OUT} = \frac{R2}{R1}\,(V_2 - V_1)$$

R1||R2=R3||R4

POPULAR ELECTRONICS ***FIG. 2-16***

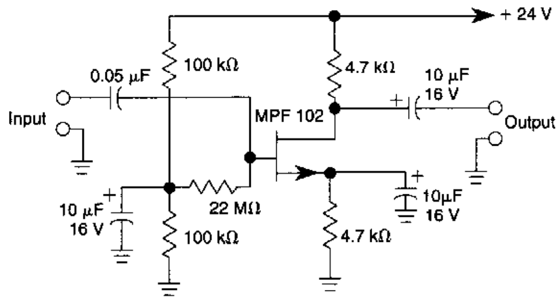
GENERAL-PURPOSE JFET PREAMP



WILLIAM SHEETS ***FIG. 2-17***

This JFET preamplifier has a gain of about 20 dB and a bandwidth of over 100 kHz. It is useful as a low-level audio amplifier for high-impedance sources.

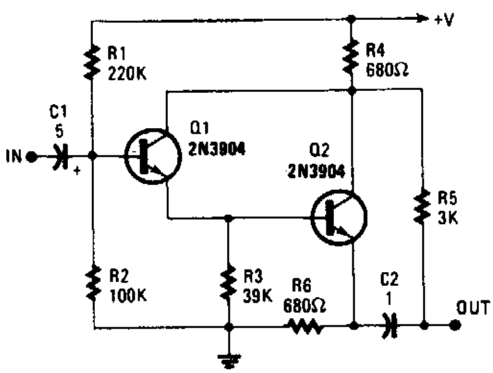
FET AMPLIFIER WITH OFFSET GATE BIAS



WILLIAM SHEETS ***FIG. 2-18***

In this amplifier circuit, the gate of the MPF102 is biased with an external voltage. This circuit achieves tighter control of the operating point and biasing conditions.

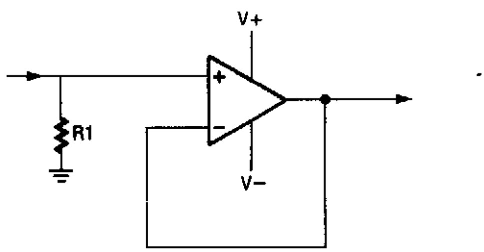
PUSH-PULL DARLINGTON AMPLIFIER



POPULAR ELECTRONICS ***FIG. 2-19***

This circuit has a high-*Z* input and push-pull output via the output taken across R4 and R6.

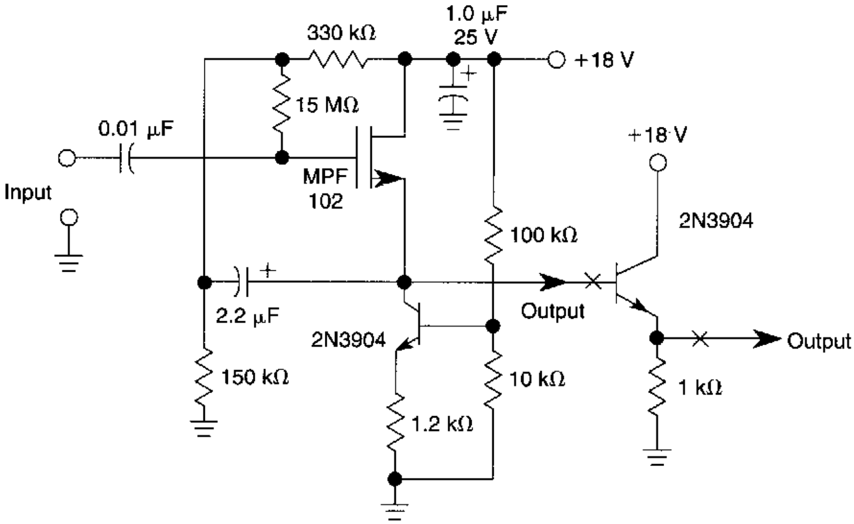
NONINVERTED UNITY-GAIN AMPLIFIER



WILLIAM SHEETS ***FIG. 2-20***

An op amp can be used as a unity gain amplifier by connecting its output to its inverting input as shown. R1 should be low enough so the bias current of the op amp does not cause an appreciable offset.

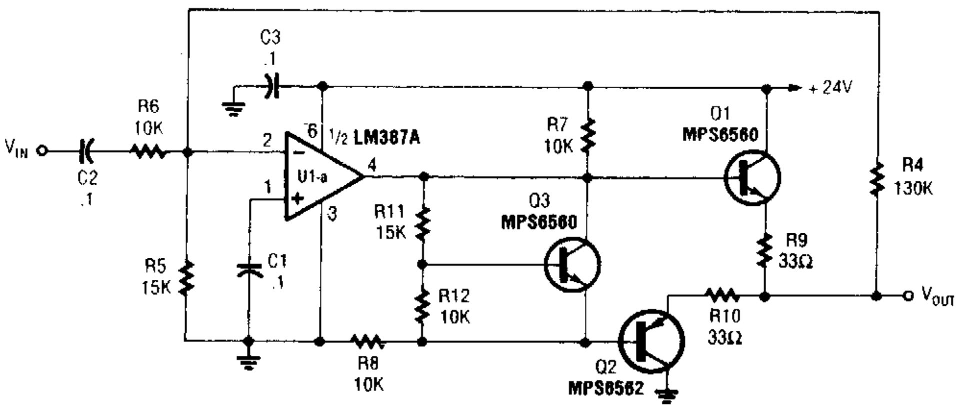
500-MΩ INPUT IMPEDANCE WITH JFET AMP



WILLIAM SHEETS *FIG. 2-21*

A current source using a 2N3904 transistor plus bootstrapping, achieves an input impedance of 500 MΩ. A second 2N3904 transistor can be added at X to lower the output impedance.

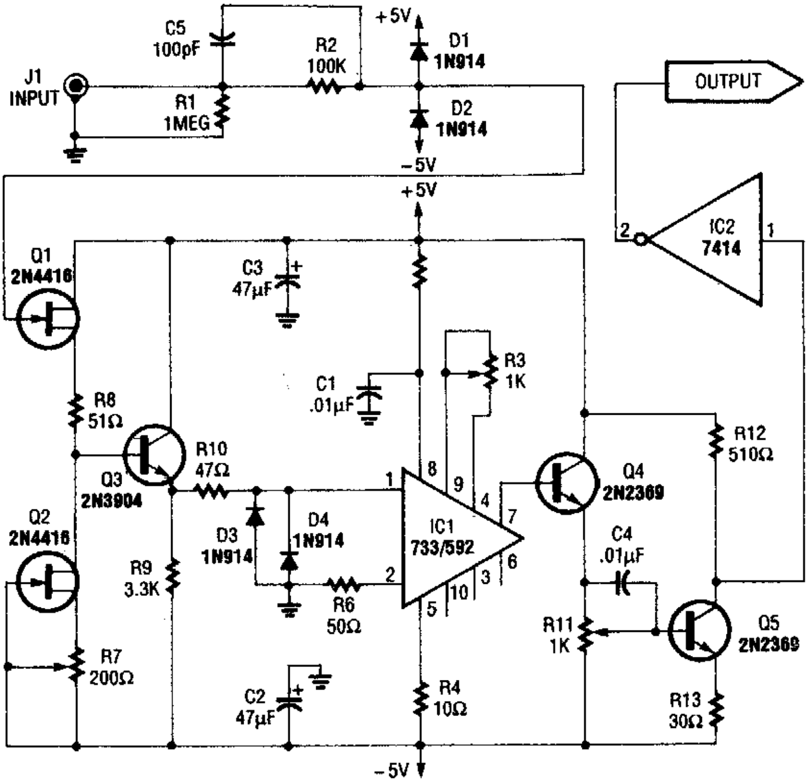
DISCRETE CURRENT-BOOSTER AMPLIFIER



POPULAR ELECTRONICS *FIG. 2-22*

Suitable as a line driver, this circuit is useable in many similar audio applications.

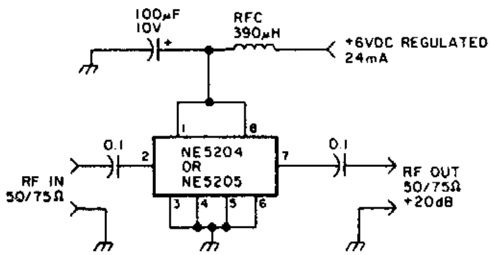
FREQUENCY COUNTER PREAMP



RADIO-ELECTRONICS ***FIG. 2-23***

Based on the LM733 or NE592, the preamp shown has a bandwidth of 100 MHz. The FET inputs provide about 1-MΩ input impedance. Q4, Q5, and IC2 provide signal conditioning.

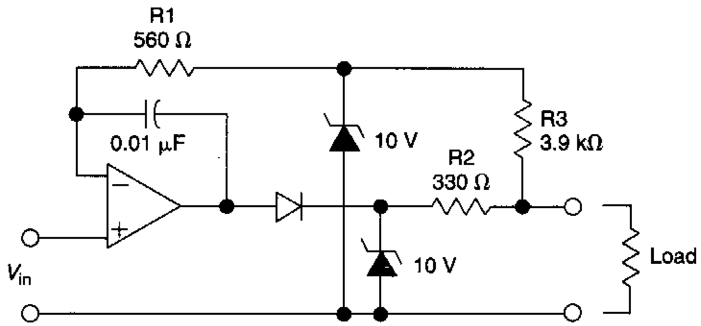
AUDIO TO UHF PREAMP



The Signetics NE5204 or NE5205 can be used in this AF to 350-MHz (–30 dB) preamp. If 600 MHz @ 3 dB is needed, use the NE5205. The noise figure is 4.8 dB at 75 Ω, 6 dB at 50 Ω. Gain is approximately +20 dB over the passband.

73 AMATEUR RADIO TODAY ***FIG. 2-24***

V- & I-PROTECTED INTRINSICALLY SAFE OP AMP



WILLIAM SHEETS

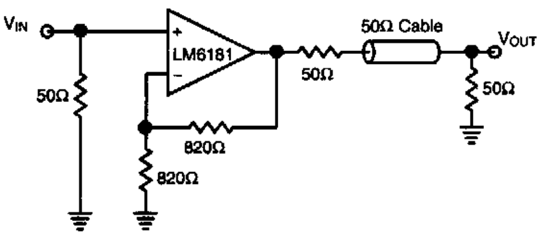
FIG. 2-25

The circuit is designed to drive an external load. A fault condition in the external load circuit could feed excessive current or voltage back into the line drive circuit. If excessive voltage appears from the load, the two zener diodes will clamp that voltage to a safe level, which in this case is 10 V.

The current in the zener diodes, op amp, and the remainder of the circuitry is limited to a safe level by resistors R1, R2, and R3. D1 protects the op-amp output stage from 10 V appearing across the clamp diodes under a fault condition.

The advantage of this circuit is that, although it's designed as unity gain buffer, the same techniques can be applied to inverting, noninverting, or differential gain stages.

CURRENT FEEDBACK AMP DELIVERS 100 mA @ 100 MHz

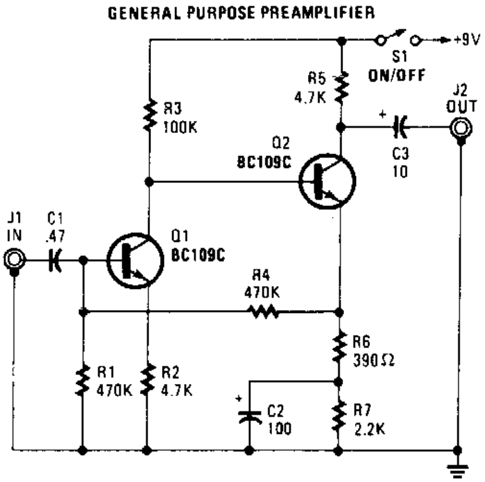


Using a NS LM6181, this IC is useful in cable drivers. The supply voltage is ±5 V to ±15 V.

NATIONAL SEMICONDUCTOR

FIG. 2-26

GENERAL-PURPOSE PREAMPLIFIER

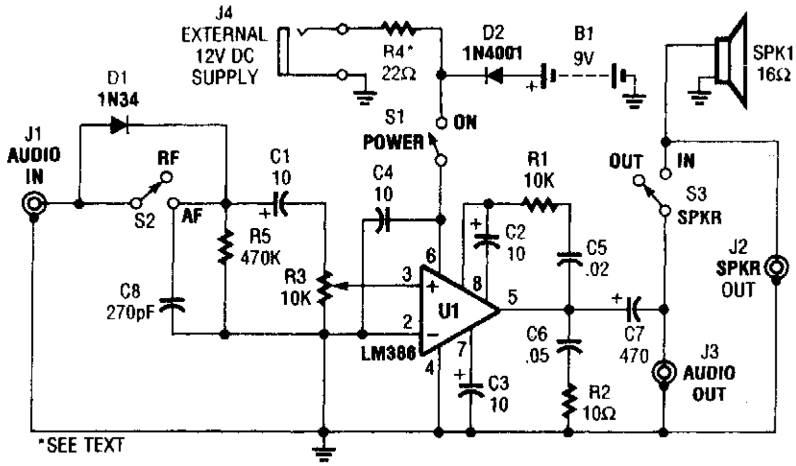


Suitable for general audio use, the preamp circuit uses a feedback pair. Current gain is set by the ratio of $(R_4 + R_6)/R_4$.

POPULAR ELECTRONICS

FIG. 2-27

TEST BENCH AMPLIFIER



POPULAR ELECTRONICS

FIG. 2-28

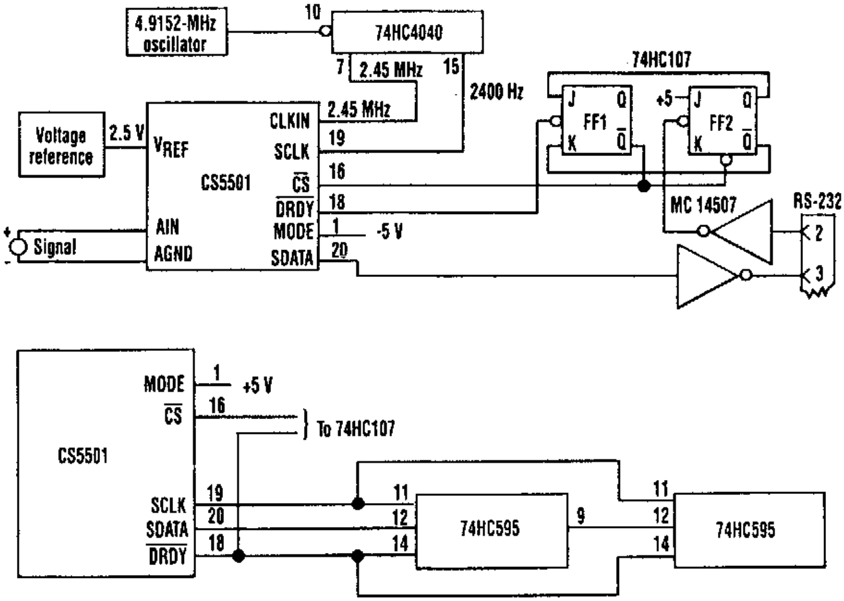
This amplifier might be useful in servicing or bench testing as a signal tracer or as a building block in various systems.

3

Analog-to-Digital Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ADC POLLER



ELECTRONIC DESIGN

FIG. 3-1

Because the CS5501 16-bit-delta-sigma analog-to-digital converter lacks a "start convert" command, it converts continuously, outputting conversion words to its output register every 1024 cycles of its master clock. However, by incorporating a standard dual J-K flip-flop into the circuit, the ADC can be configured to output a single-conversion word only when it is polled.

The CS5501 converter can be operated in its asynchronous communication mode (UART) to transmit one 16-bit conversion word when it is polled over an RS-232 serial line (see figure). A null character (all zeros) is transmitted to the circuit and sets the flip-flop FF2. The CS5501 can then output a single-conversion word, which is transmitted over the RS-232 line as two bytes with start and stop bits.

The baud rate can be chosen by selecting the appropriate clock divider rate on the 74HC4040 counter/divider as the serial port clock (SLCK) for the ADC. This type of polled-mode operation is also useful when the ADC's output register is configured to operate in the synchronous-serial clock (SSC) mode. In this case, the converter will load one output word into a 16-bit serial-to-parallel register (two 74HC595 8-bit registers) when polled to do so (see figure).

8-CHANNEL A/D CONVERTER FOR PC CLONES

The following program causes the A-D converter to perform eight sequential conversions and display the result. It's written in Turbo BASIC/Power BASIC source code, but it will run under the GW-BASIC interpreter if you replace the delay statements with FOR/NEXT loops, and add line numbers as shown in the second listing. These programs are available on the 73 BBS under the filenames ADC Turbo.BAS and ADCGW.BAS.

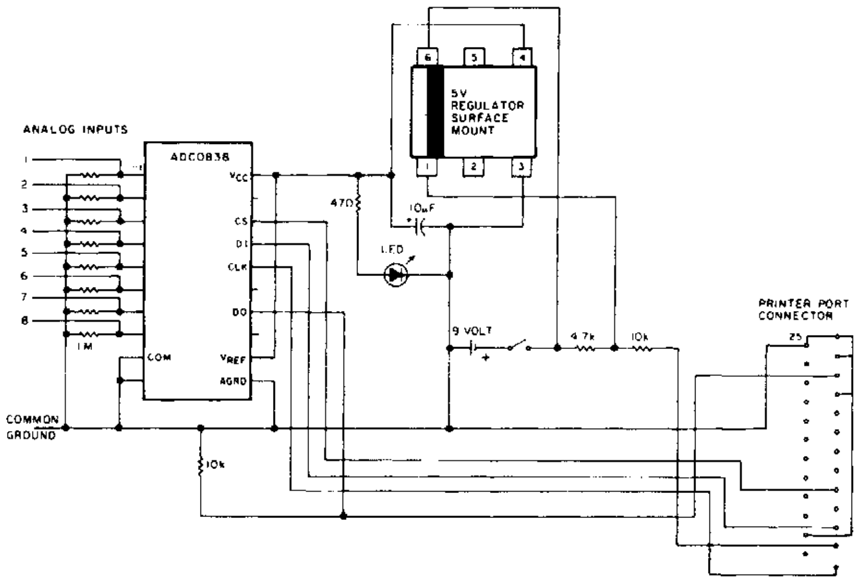
```
INITIALIZE:  'remarks follow the apostrophe
  screen 0                          'text mode 80 columns
  color 14,0                        'yellow on blue
  cls                               'clear the screen
  clear                             'clear all variables
  toggle%=2                         'initialize variables
  oddsign%=0

MINORLOOP:
  while not instat                  'keep going until a key is pressed
    out 888,1                       'regulator line high
  delay 1                           'wait 1 second before next sample
    out 888,0                       'light up the regulator
    delay .054                      'wait 54 milliseconds to stabilize
  for ch%=0 to 7                    'scan 8 channels
    out 888,8                       'CS high pin 5
    out 888,0                       'CS low
    out 888,2                       'start bit is always high DI line
    out 890,0                       'clock high pin 1 of DB 25 printer
      for slow%=0 to 1:next slow%   'stretches clock pulse
    out 890,1                       'clock low
    out 888,2                       '8 single ended measurements selected
    out 890,0                       'clock high
      for slow%=0 to 1:next slow%   'stretches clock pulse
    out 890,1                       'clock low
    out 888,oddsign%                'part of the channel selection string
    swap oddsign%,toggle%           'toggles between high and low
    out 890,0                       'clock high
      for slow%=0 to 1:next slow%   'stretches clock pulse
    out 890,1                       'clock low
    out 888,select1%                'part of the channel selection string
    out 890,0                       'clock high
      for slow%=0 to 1:next slow%   'stretches clock pulse
    out 890,1                       'clock low
    out 888,select0%                'part of the channel selection string
    out 890,0                       'clock high
      for slow%=0 to 1:next slow%   'stretches clock pulse
  out 890,1                         'clock low
READBITS:
    for bit%=7 to 0 step -1         'MSB is first out
    out 890,0                       'clock high
      for slow%=0 to 1:next slow%   stretches clock pulse
    out 890,1                       'clock low
    ad%=inp(889)                    'port 889 pin 10  7-low135-high
    if ad%<120 then byte%=byte%+(2^bit%)
    next bit%
    if ch%=0  then  select1%=0 : select0%=0 : ch0volts=byte%/51
    if ch%=1  then  select1%=0 : select0%=2 : ch1volts=byte%/51
    if ch%=2  then  select1%=0 : select0%=2 : ch2volts=byte%/51
    if ch%=3  then  select1%=2 : select0%=0 : ch3volts=byte%/51
    if ch%=4  then  select1%=2 : select0%=0 : ch4volts=byte%/51
    if ch%=5  then  select1%=2 : select0%=2 : ch5volts=byte%/51
    if ch%=6  then  select1%=2 : select0%=2 : ch6volts=byte%/51
    if ch%=7  then  select1%=0 : select0%=0 : ch7volts=byte%/51
    byte%=0
  next ch%
  print using
  "##.#";ch0volts, ch1volts, ch2volts,ch3colts,ch4volts,ch5volts,ch6volts,ch7volts
  wend
```

GWBASIC Version

```
10 ' The following program causes the A-D converter to perform eight
20 ' sequential conversions and display the result.
30 SCREEN 0                               'text mode 80 columns
40 COLOR 14,0                             'yellow on blue
50 CLS                                    'clear the screen
60 CLEAR                                  'clear all variables
70 TOGGLE%=2                              'initialize variables
80 ODDSIGN%=0
90 IF INKEY$<>"" THEN END                 'keep going until a key is pressed
100 OUT 888,1                             'regulator line high
110 OUT 888,0                             'light up the regulator
120 FOR W%=0 TO 500:NEXT W%               'wait 54 mseconds to stabilize
130 FOR CH%=0 TO 7                        'scan 8 channels
140 OUT 888,8                             'CS high  pin 5
150 OUT 888,0                             'CS low
160 OUT 888,2                             'start bit is always high DI line
170 OUT 890,0                             'clock high pin 1 of DB 25 printer
180 FOR SLOW%=0 TO 1:NEXT SLOW%           'stretches clock pulse
190 OUT 890,1                             'clock low
200 OUT 888,2                             '8 single ended measurements selected
210 OUT 890,0                             'clock high
220 FOR SLOW%=0 TO 1:NEXT SLOW%           'stretches clock pulse
230 OUT 890,1                             'clock low
240 OUT 888,ODDSIGN%                      'part of the channel selection string
250 SWAP ODDSIGN%,TOGGLE%                 'toggles between high and low
260 OUT 890,0                             'clock high
270 FOR SLOW%=0 TO 1:NEXT SLOW%           'stretches clock pulse
280 OUT 890,1                             'clock low
290 OUT 888,SELECT1%                      'part of the channel selection string
300 OUT 890,0                             'clock high
310 FOR SLOW%=0 TO 1:NEXT SLOW%           'stretches clock pulse
320 OUT 890,1                             'clock low
330 OUT 888,SELECT0%                      'part of the channel selection string
340 OUT 890,0                             'clock high
350 FOR SLOW%=0 TO 1:NEXT SLOW%           'stretches clock pulse
360 OUT 890,1                             'clock low
370 REM                                   'read output bits
380 FOR BIT%=7 TO 0 STEP -1               'MSB is first out
390 OUT 890,0                             'clock high
400 FOR SLOW%=0 TO 1:NEXT SLOW%           if ad%'stretches clock pulse
410 OUT 890,1                             'clock low
420 AD%=INP(889)                          'port 889 pin 10  7-low  135-high
430 IF AD%<120 THEN BYTE%=BYTE%+(2^BIT%)
440 NEXT BIT%
450 IF  CH%=0  THEN  SELECT1%=0 : SELECT0%=0 : CH0VOLTS=BYTE%/51
460 IF  CH%=1  THEN  SELECT1%=0 : SELECT0%=2 : CH1VOLTS=BYTE%/51
470 IF  CH%=2  THEN  SELECT1%=0 : SELECT0%=2 : CH2VOLTS=BYTE%/51
480 IF  CH%=3  THEN  SELECT1%=2 : SELECT0%=0 : CH3VOLTS=BYTE%/51
490 IF  CH%=4  THEN  SELECT1%=2 : SELECT0%=0 : CH4VOLTS=BYTE%/51
500 IF  CH%=5  THEN  SELECT1%=2 : SELECT0%=2 : CH5VOLTS=BYTE%/51
510IF  CH%=6  THEN  SELECT1%=2 : SELECT0%=2 : CH6VOLTS=BYTE%/51
520IF  CH%=7  THEN  SELECT1%=0 : SELECT0%=0 : CH7VOLTS=BYTE%/51
530 BYTE%=0
540 next ch%
550 PRINT USING"##.#";CH0VOLTS, CH1VOLTS, CH2VOLTS,CH3COLTS,
    CH4VOLTS,CH5VOLTS,CH6VOLTS,CH7VOLTS
```

FIG. 3-2



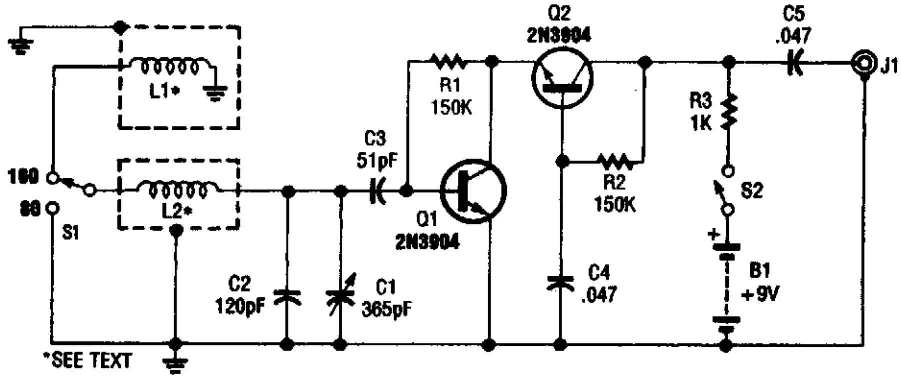
An A/D converter by National Semiconductor (ADC0838), converts 0- to 5-V analog inputs to a digital data format. A 9-V battery is used. The converter connects to the pointer port connector via a 25-pin connector.

4

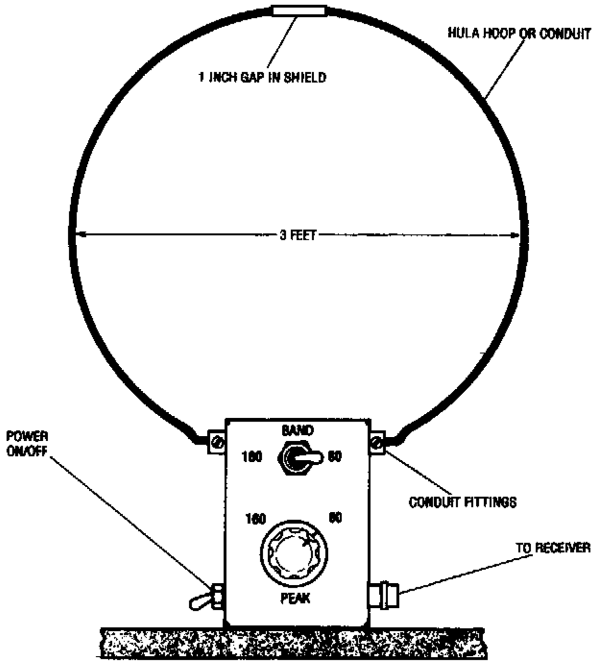
Antenna Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

DUAL-BAND LOOP ANTENNA FOR 80 & 160 m



A



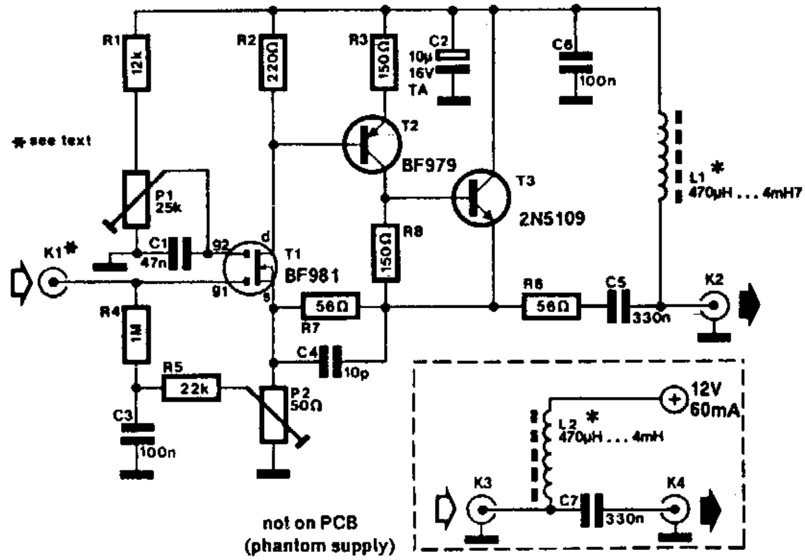
B

POPULAR ELECTRONICS

FIG. 4-1

This antenna might help to reduce power-line noise. A plastic "hula hoop" or conduit 3 feet in diameter, covered with aluminum foil as a shield is used for L1 and L2. L1 is two turns and L2 is one turn, threaded through the loop. S1 selects 160- or 80-m operation. Q1 and Q2 form a preamplifier for the loop antenna. Do not transmit with this antenna—it is for receiving only.

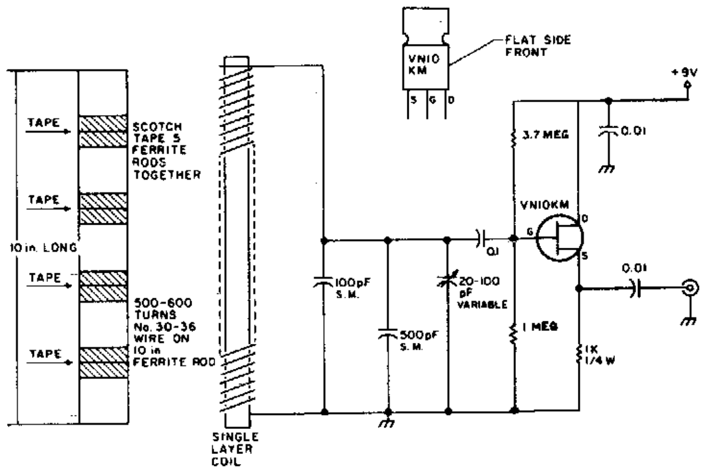
VLF/VHF WIDEBAND LOW-NOISE ACTIVE ANTENNA



ELECTOR ELECTRONICS USA ***FIG. 4-2***

A 30- to 50-cm whip antenna provides reception from 10 kHz to over 220 MHz. T1, a dual-gate MOSFET, provides low noise, high-input impedance, and high gain. The circuit is powered via the coaxial cable used to connect the antenna to a receiver.

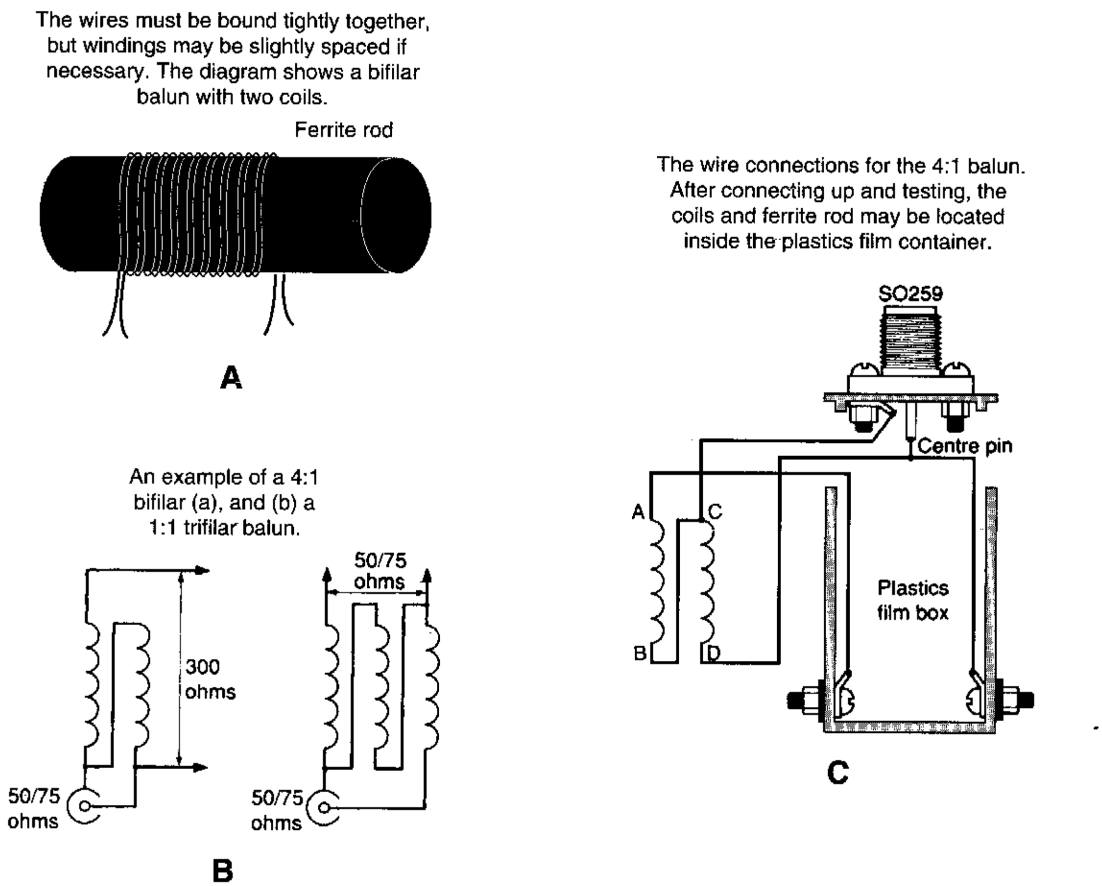
VLF 60-kHz ANTENNA/PREAMP



73 AMATEUR RADIO TODAY ***FIG. 4-3***

Suitable for 60-kHz standard frequency reception, here is a schematic for a FET preamp and antenna.

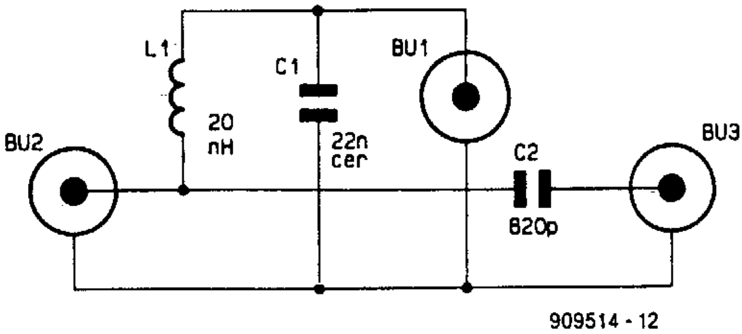
SIMPLE BALUN



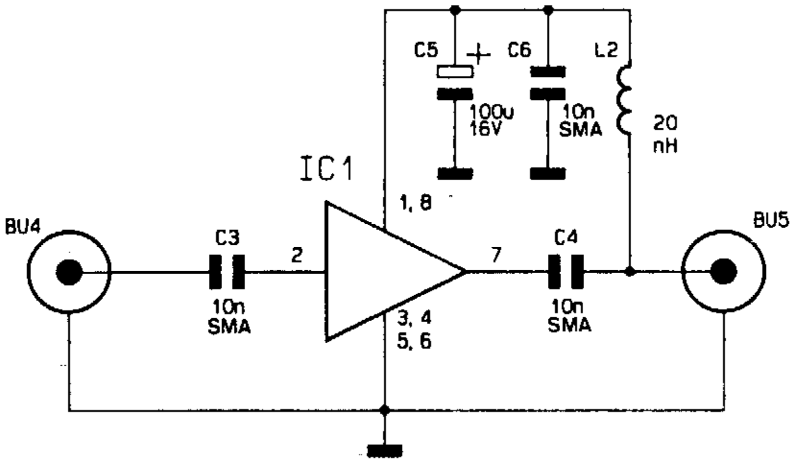
PRACTICAL WIRELESS ***FIG. 4-4***

An old ferrite rod from a junked broadcast receiver can be used to construct an antenna balun, as shown.

WIDEBAND ANTENNA PREAMPLIFIER



A

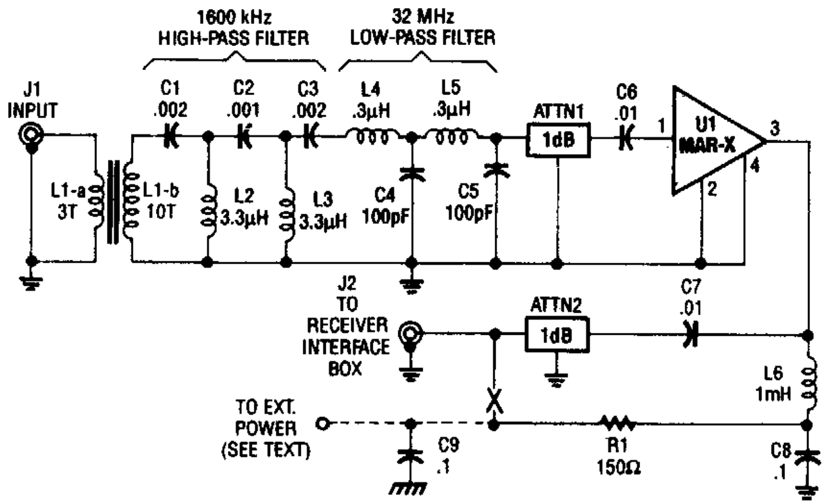


B

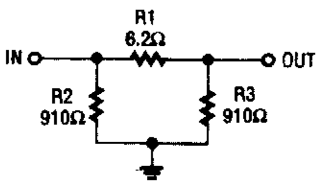
ELEKTOR ELECTRONICS ***FIG. 4-5***

This wideband antenna preamplifier has a gain of around 20 dB from 40 to 860 MHz, covering the entire VHF, FM, commercial, and UHF bands. A phantom power supply provides dc to the preamp via the coaxial cable feeding the unit.

HF BROADBAND ANTENNA PREAMP



A



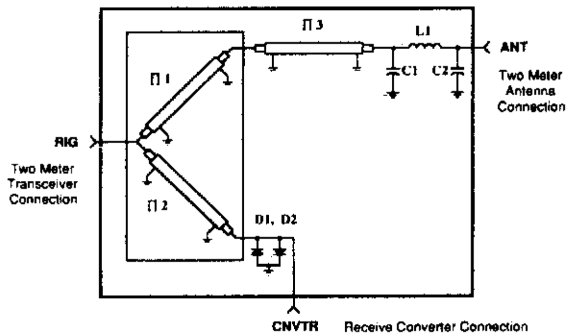
B

POPULAR ELECTRONICS *FIG. 4-6*

The HF/SW receiver preamplifier is comprised of a broadband toroidal transformer (L1-a and L1-b), LC network (comprised of a 1600-kHz, high-pass filter and a 32-MHz, low-pass filter), L2 and L3 (26 turns of #26 enameled wire wound on an Amidon Associates T-50-2, red, toroidal core), a pair of resistive attenuators (ATTN1 and ATTN2), and a MAR-x device.

Shown here is the composition of a basic 1-dB pi-network resistor antenuator. This is the method of supplying dc power to a preamplifier using only the RF coax cable.

AUTOMATIC TR SWITCH



Indicates Wilkinson Hybrid section - See text for discussion

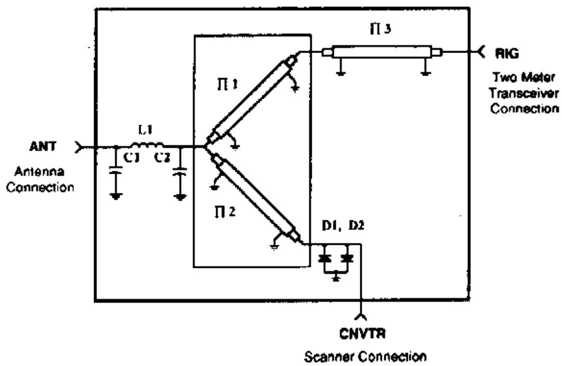
C1, C2 - 39pF mica caps

D1, D2 - 1N914, 1N4148 Si Diodes

L1 - 2 turns # 18 tinned wire, 1/4 inch ID, 0.2 inch long

Π1, Π2 and **Π3** consist of 75 Ω coax sections, 1/4 wave at the center of the transceiver transmitter band typically 147 Mhz.
Π1 and **Π3** are combined in one continuous length of cable - 1/2 wavelength total.
See text for additional discussion.

A



C1, C2 - 39pF mica caps

D1, D2 - 1N914, 1N4148 Si Diodes

L1 - 2 turns # 18 tinned wire, 1/4 inch ID, 0.2 inch long

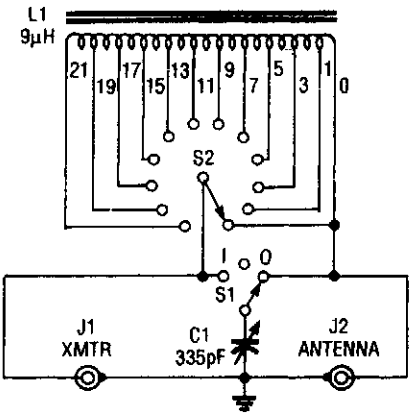
Π1, Π2 and **Π3** consist of 75 Ω coax sections, 1/4 wave at the center of the transceiver transmitter band typically 147 Mhz.
Π1 and **Π3** are combined in one continuous length of cable - 1/2 wavelength total.
See text for additional discussion.

B

73 AMATEUR RADIO TODAY ***FIG. 4-7***

A pair of diodes and a quarter-wave transmission line are used as an automatic TR switch. D1 and D2 conduct during transmit periods, short-circuiting the scanner input. In this mode, the ¼-wave line appears as an open circuit. In receive, the circuit acts as a Wilkinson power divider.

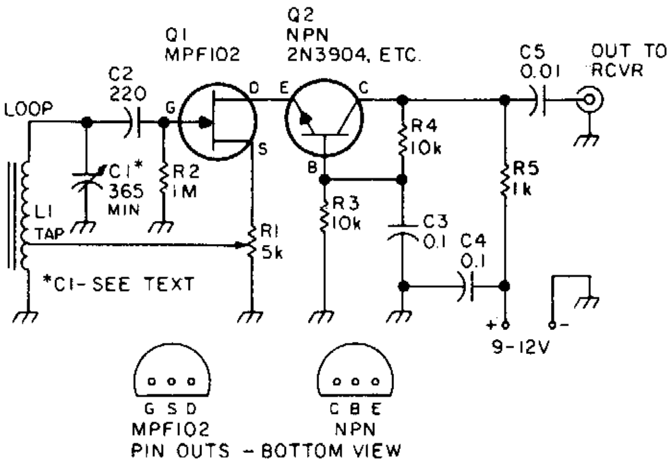
LOW-POWER ANTENNA TUNER



This antenna tuner is suitable for use with low-power (less than 5 W) transmitters or SW receivers. S2 selects inductance and S2 connects the 365-pF capacitor to either the transmitter or the side of the inductor. The tiny tuner is comprised of a tapped inductor (L1) and a variable capacitor (C1), which is connected to the inductor through a center-off SPDT switch (S1). That switch arrangement permits the capacitor to be connected to either the input or the output of the circuit.

1993 ELECTRONICS HOBBYISTS HANDBOOK ***FIG. 4-8***

LOOP ANTENNA PREAMPLIFIER



73 AMATEUR RADIO TODAY ***FIG. 4-9***

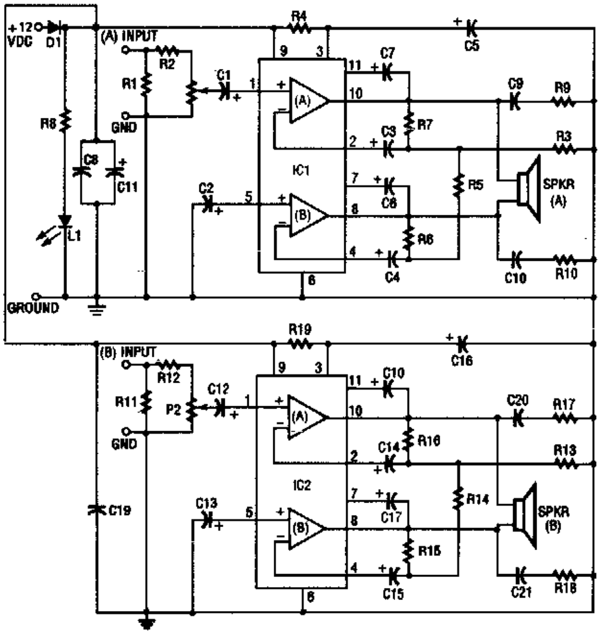
This preamplifier has a built-in regeneration control boost gain selectivity. C1 is a single or multi-gang AM broadcast-band tuning capacitor. L1 is a ferrite loop antenna, tapped at about 15 to 25% of total turns. This circuit should prove useful for low-frequency (up to 3 MHz) reception, where a loop would be advantageous to reduce man-made noise pickup.

5

Audio Power Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

20-W + 20-W STEREO AMPLIFIER



C1,C2,C12,C13 ...... 4.7 µF Electrolytic Capacitor
C3,C4,C5,C6,
C7,C14,C15 .. 100 µF Electrolytic Capacitor
C16,C17,C18,C8,
C9,C10,C19,C20,
C21 ...... 0.47 µF Mylar Capacitor
C11 ...... 3300 µF, 25V Electrolytic Capacitor
D1 .... 3 Amp Rectifier
L1 .... Red LED
R1, R11 .. 47 ohm, 1 watt Resistor
R2, R12 .. 33K, 1/4 watt Resistor
R3,R5,R13,R14 ...... 12 ohm, 1/4 watt Resistor
R4,R19 .. 120K, 1/4 watt Resistor
R6,R15 .. 2K. 1/4 watt Resistor
R7,R16 .. 1K, 1/4 watt Resistor
R8 .... 680 ohm, 1/4 watt Resistor
R9,R10,R17,R18 .... 3 ohm, 1/4 watt Resistor
P1,P2 .... 50K Dual Potentiometer
IC1, IC2 NEC70001AB Amplifier

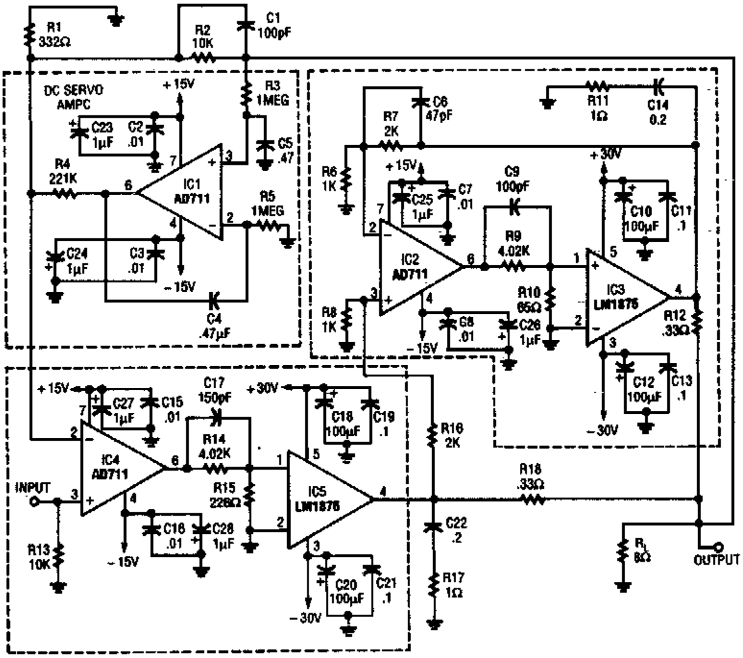
1991 PE HOBBYIST HANDBOOK ***FIG. 5-1***

The 20-W + 20-W stereo amp consists of two complete, separate 20-W RMS bridge-type amplifiers. The input signal source is brought into the amplifier through the voltage divider network, which is made up of R1, R2, and P1. Resistor R1 provides a load impedance between the signal source and ground. Resistor R2 couples that signal to potentiometer P1.

The signal is coupled by capacitor C1 to the noninverting (+) input (pin 1) of internal amplifier (A) of IC1, where the signal is greatly amplified. Capacitor C2 couples the (+) input of the other (B) internal amplifier of IC1 to ground. That causes the input signal, which is referenced to ground, to be coupled to both amplifiers because both the inputs and outputs of IC1 (A) and IC1 (B) are connected in a bridge configuration. Notice that the output of IC1 (A) from pin 10 is connected to one side of the speaker and the output of IC1 (B) from pin 8 is connected to the other side of the speaker. That is why the speakers used cannot have one side connected to ground. Resistors R6 and R7 set the gain of the amplifier. Resistors R9 and R10 and capacitors C9 and C10 provide frequency stability and prevent oscillation. Capacitors C6 and C7 provide "bootstrapping," which prevents distortion at low frequencies. LED L1 lights up by way of a series resistor connected from the anode to +12 Vdc when power is applied.

Power for both IC1 and IC2 is brought in through D1 (to protect amplifiers from reverse polarity). Capacitor C11 provides additional power supply line filtering. This booster is capable of producing 20 W RMS output out of each channel.

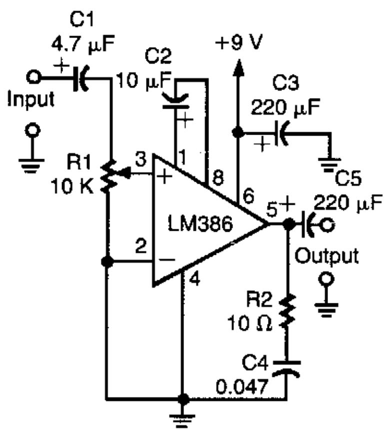
40-W AMPLIFIER



ELECTRONICS NOW *FIG. 5-2*

This circuit uses two LM1875 devices and a dc servo loop. This circuit provides 40-W output. IC3 and IC5 must be heatsinked.

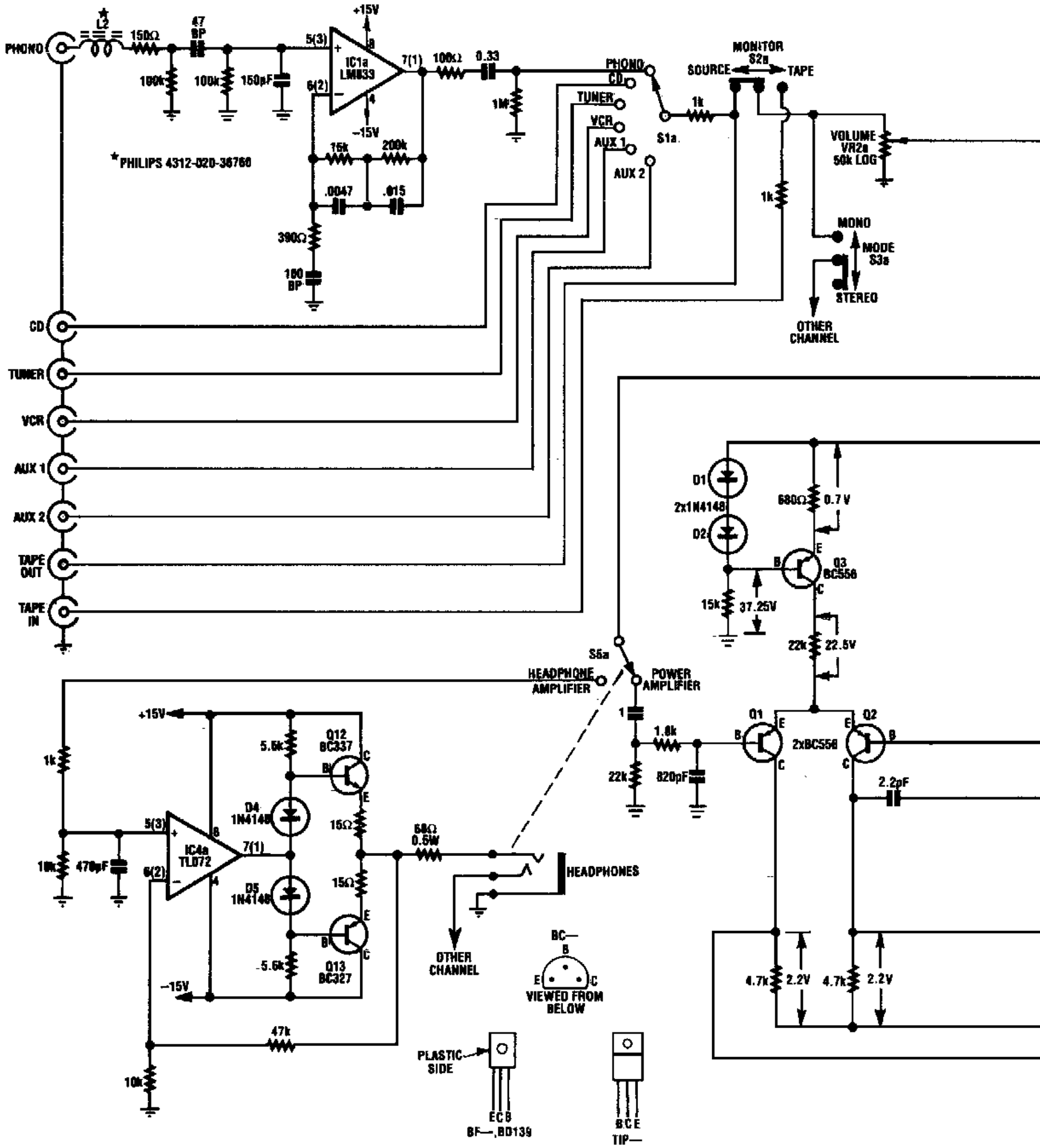
HALF-WATT SINGLE-CHANNEL AUDIO AMPLIFIER



This circuit uses an LM386 IC and will work from 6- to 12-V battery sources. Output is about 0.5 W into 8 Ω.

ELECTRONICS NOW *FIG. 5-3*

DUAL AUDIO AMPLIFIER



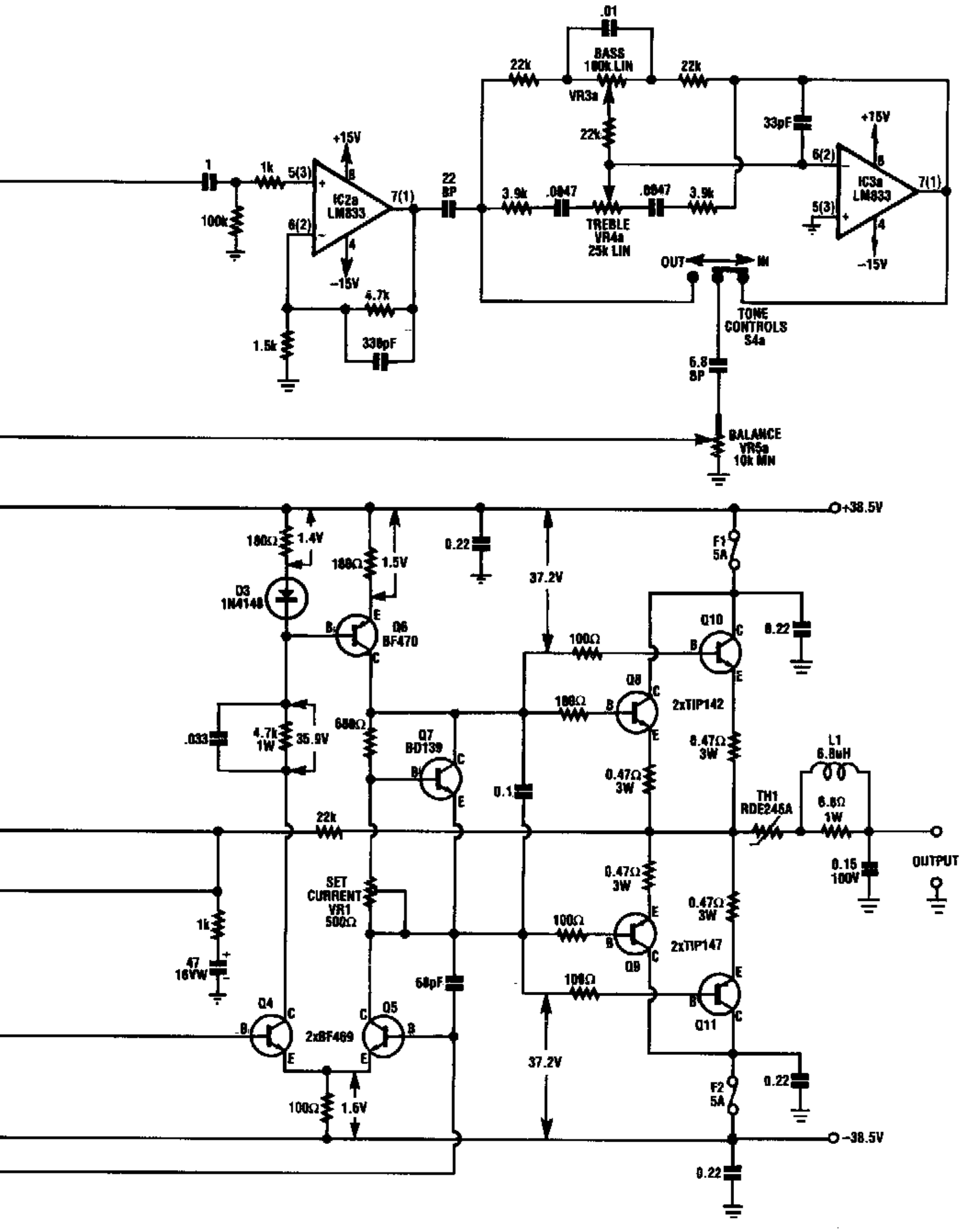
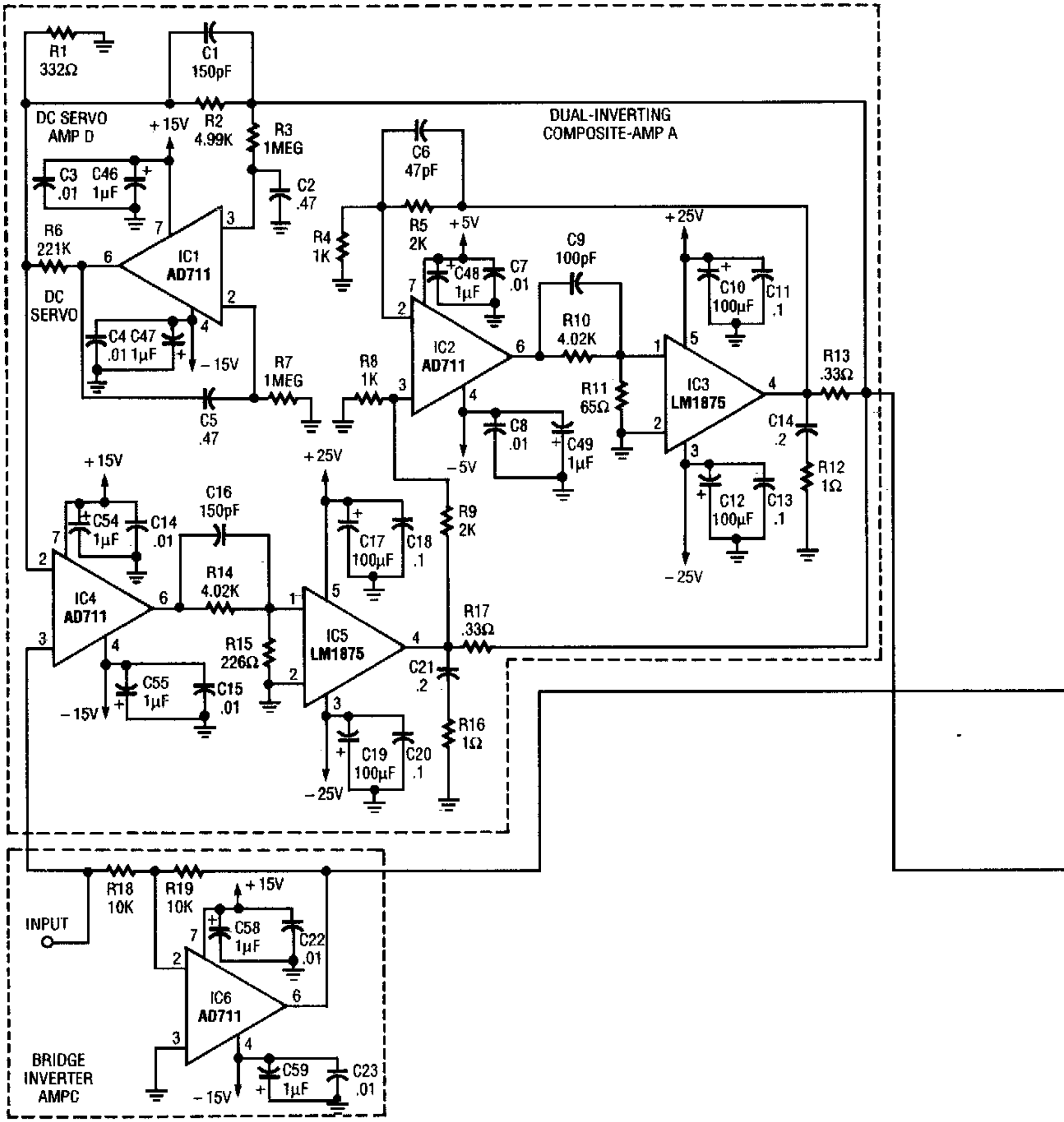
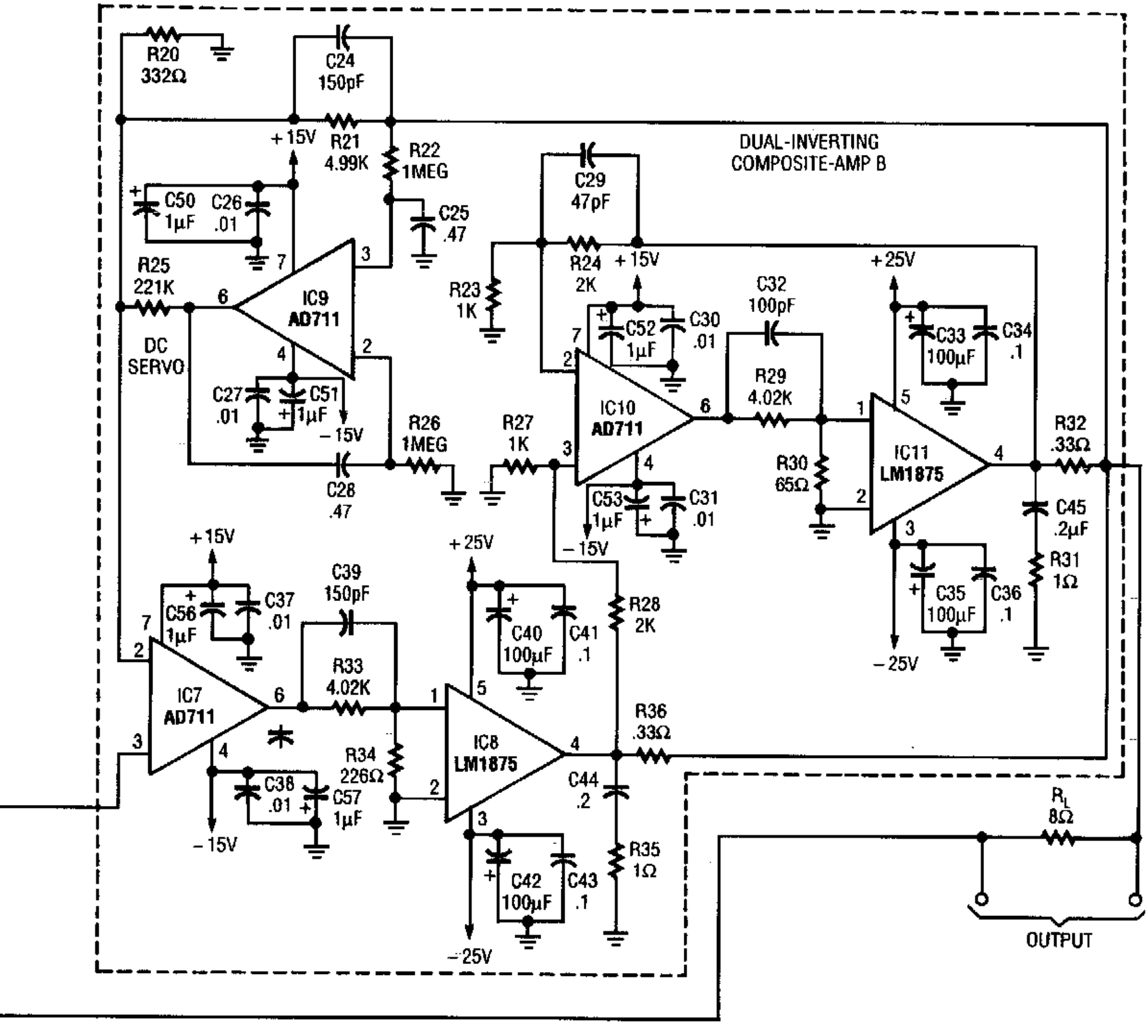

.01
BASS
100k LIN
VR3a
22k
22k
22k
33pF
+15V
−15V
1
1k
5(3)
100k
6(2)
IC2a
LM833
7(1)
22
BP
3.9k
.0047
.0047
3.9k
TREBLE
VR4a
25k LIN
IC3a
LM833
OUT
IN
4.7k
1.5k
330pF
TONE
CONTROLS
S4a
6.8
BP
BALANCE
VR5a
10k MN
+38.5V
180Ω
1.4V
180Ω
1.5V
0.22
37.2V
F1
5A
D3
1N4148
Q6
BF470
Q10
100Ω
Q8
2xTIP142
.033
4.7k
1W
35.9V
680Ω
Q7
BD139
0.47Ω
3W
L1
6.8μH
0.1
TH1
RDE245A
6.8Ω
1W
OUTPUT
0.15
100V
SET
CURRENT
VR1
500Ω
1k
47
16VW
Q9
2xTIP147
Q11
68pF
Q4
Q5
2xBF469
F2
5A
1.6V
−38.5V

FIG. 5-4

A 70-W COMPOSITE AMPLIFIER

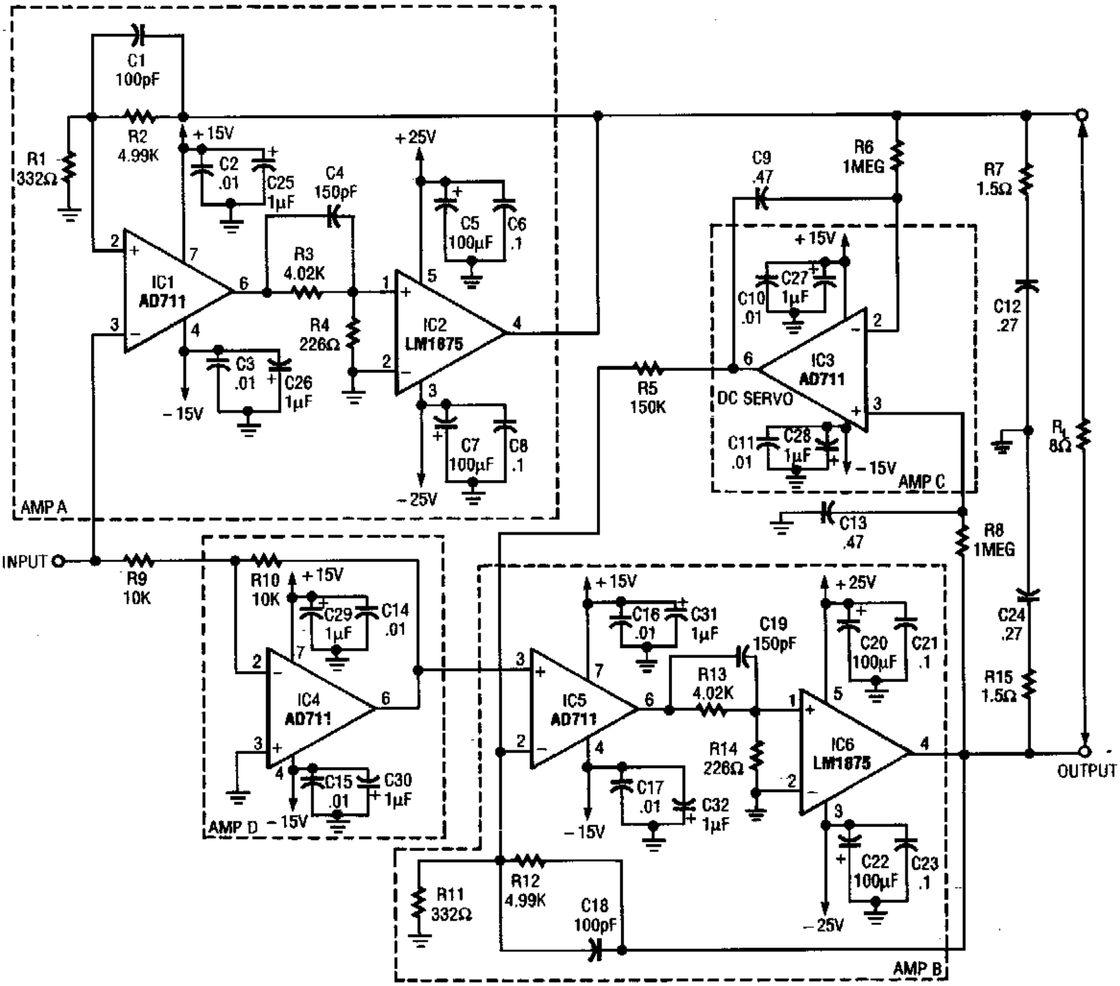




Four LM1875 devices, suitably heatsinked, and a ±25-V supply, 70 W of output are available from this circuit. IC6 is a phase inverter.

FIG. 5-5

A 33-W BRIDGE COMPOSITE AMPLIFIER

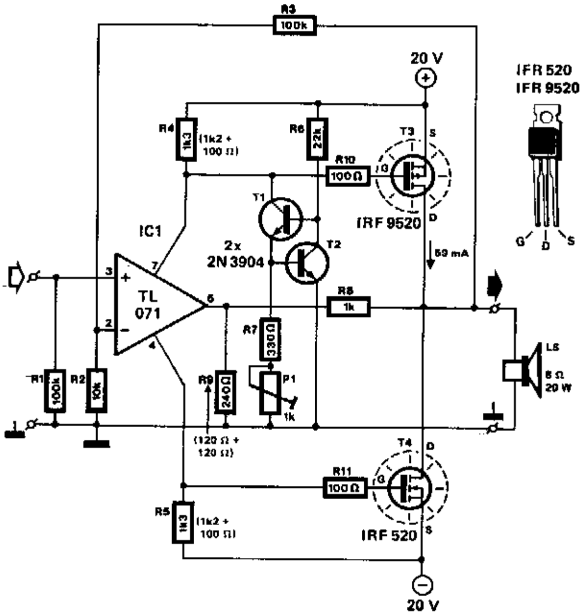


ELECTRONICS NOW

FIG. 5-6

Two LM1875 ICs provide 33 W of audio. IC4 is used as a phase inverter. IC6 and IC2 must be heatsinked.

MOSFET POWER AMPLIFIER

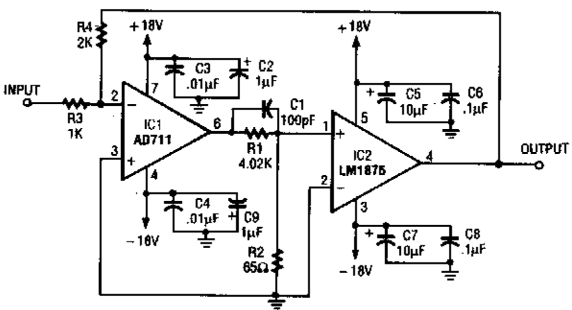


303 CIRCUITS

FIG. 5-7

Two complementary MOSFETs are used to deliver 20 W into 8 Ω. A TL071 op amp is used as an input amplifier. The MOSFETs should be heatsinked with a heatsink of better than 5°C/W capability. THD is less than 0.15% from 100 Hz to 10 kHz.

10-W NONINVERTING COMPOSITE AMPLIFIER

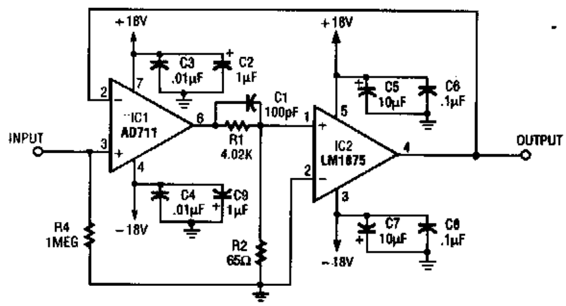


ELECTRONICS NOW

FIG. 5-8

By using an LM1875, suitably heatsinked, a 10-W amplifier that uses two IC devices can be built. IC2 must be heatsinked.

10-W INVERTING COMPOSITE AMPLIFIER

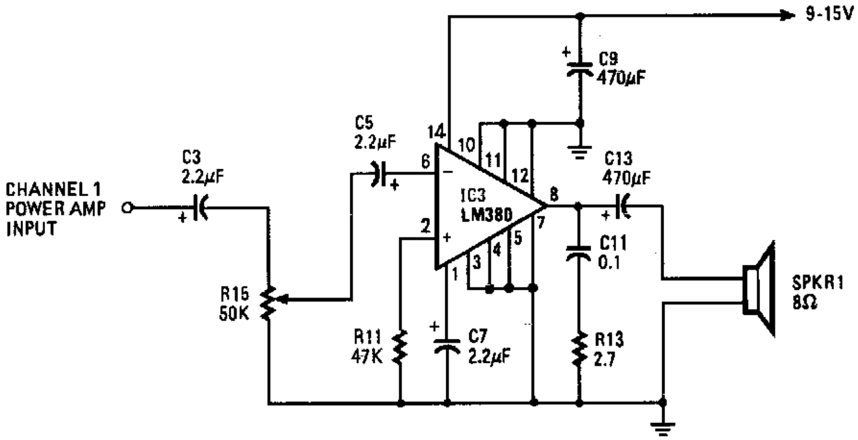


ELECTRONICS NOW

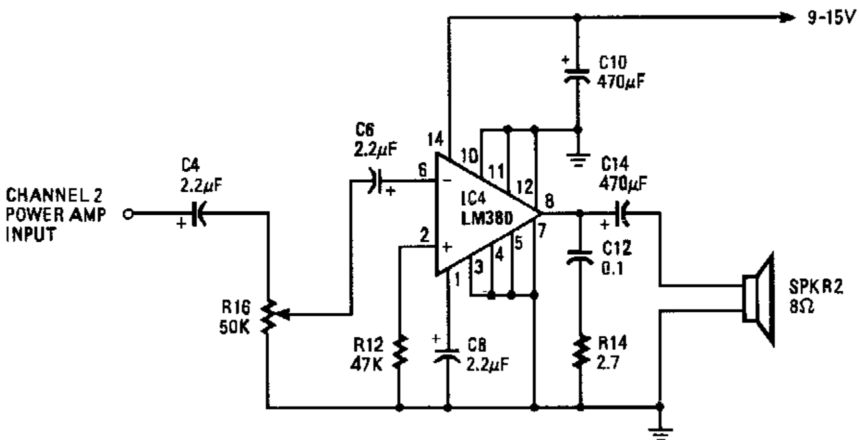
FIG. 5-9

Using an LM1875, a 10-W amplifier can be build using just two IC devices. The gain = R_4/R_3. Note that IC12 must be heatsinked.

LM380 PERSONAL STEREO AMPLIFIER



A

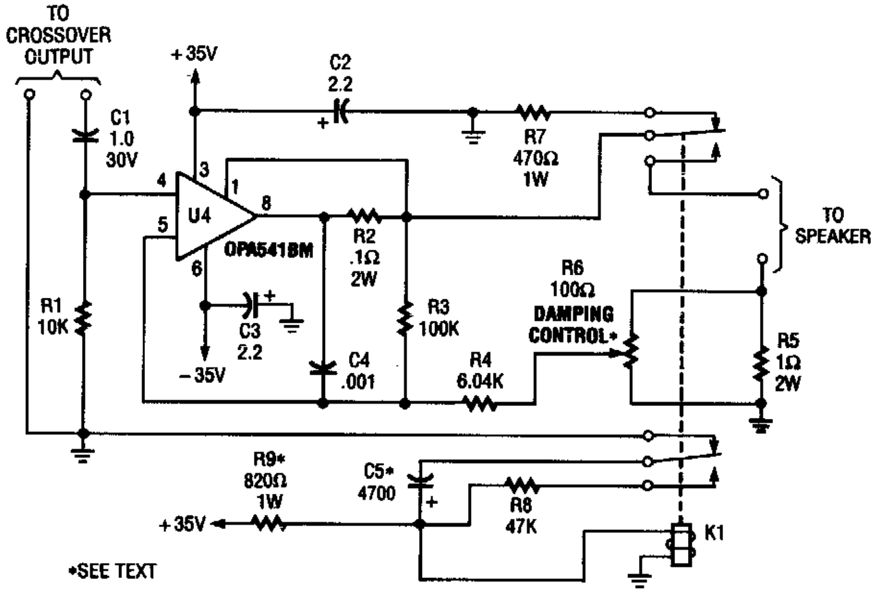


B

1987 R-E EXPERIMENTERS HANDBOOK *FIG. 5-10*

With the simple circuit, you can use your personal stereo to drive standard 8-Ω speakers.

SUBWOOFER AMPLIFIER

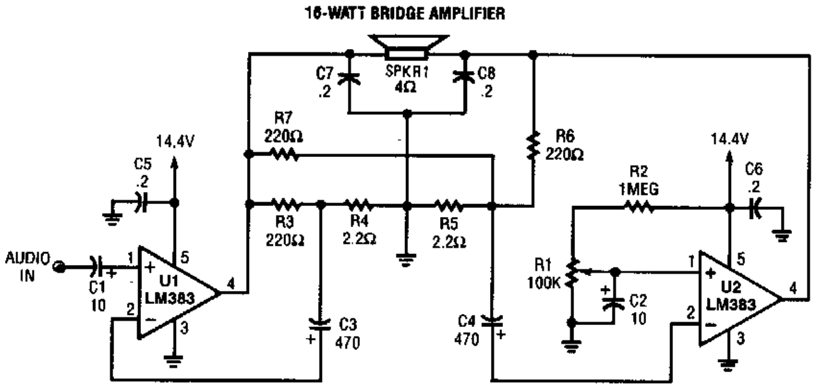


POPULAR ELECTRONICS

FIG. 5-11

Designed to feed a low-frequency subwoofer speaker system, the amplifier is capable of up to 100 W into an 8-Ω load. The OPA541BM op amp requires heatsinking and is manufactured by Burr-Brown Corporation. A damping control and a relay to eliminate turn-on and turn-off thump in the speaker is included.

18-W BRIDGE AUDIO AMPLIFIER

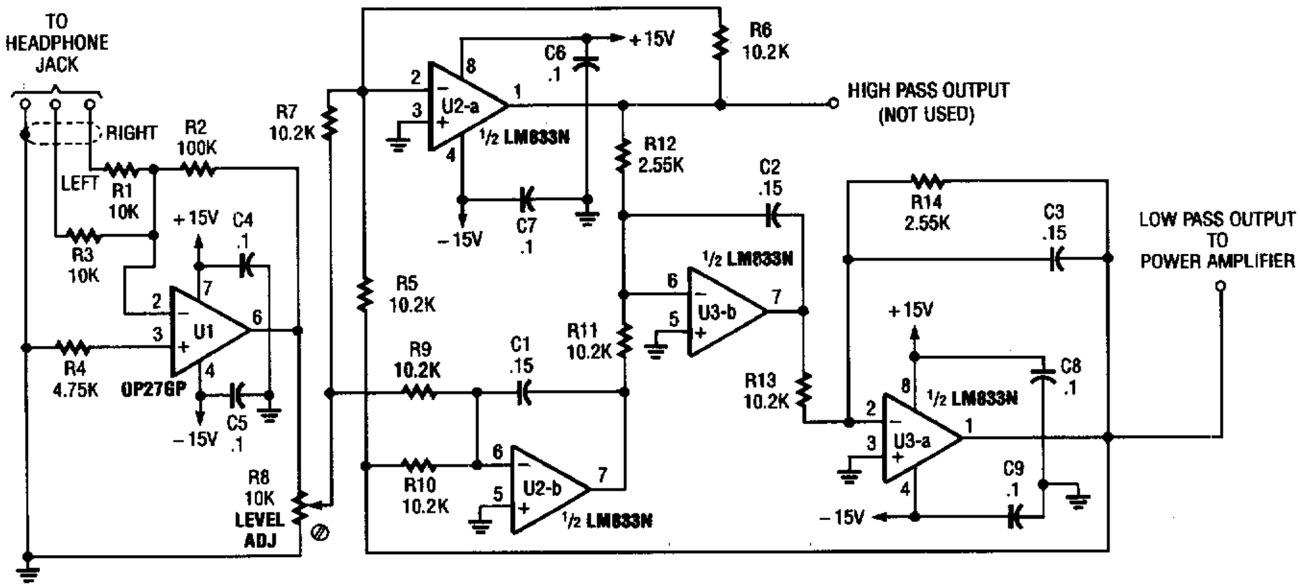


POPULAR ELECTRONICS

FIG. 5-12

Two LM383 IC devices are used in a bridge circuit that is useful for auto sound applications.

SUBWOOFER CROSSOVER AMPLIFIER

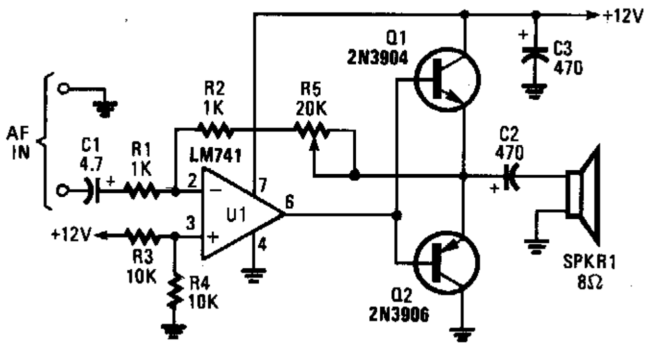


RADIO-ELECTRONICS

FIG. 5-13

The electronic-crossover circuit contains a summing amplifier that combines the left and right channels from a stereo's headphone jack. Originally used in a subwoofer system, the above circuit might be useful in similar audio applications.

AUDIO POWER AMPLIFIER

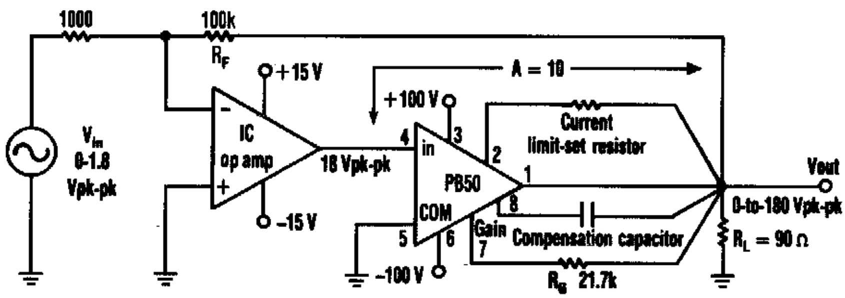


POPULAR ELECTRONICS

FIG. 5-14

The circuit, built around an LM741 op amp configured as an inverting amplifier, is used to drive complementary transistors (Q1 and Q2). The op amp's feedback loop includes the base-emitter junctions of both transistors—an arrangement that helps to reduce crossover distortion that would normally occur as a result of the emitter-to-base junction voltage drop of about 0.6 V. Potentiometer R5 varies the amplifier's voltage gain from 1 to about 20. As much as 0.5 W can be obtained from the circuit if a heatsink is added to the transistors.

FAST HIGH-VOLTAGE LINEAR POWER AMP



ELECTRONIC DESIGN

FIG. 5-15

An Apex PB50 Booster Amplifier, plus an IC op amp, can be used in a high-voltage op amp that converts a small analog signal to a 180-V p-p signal.

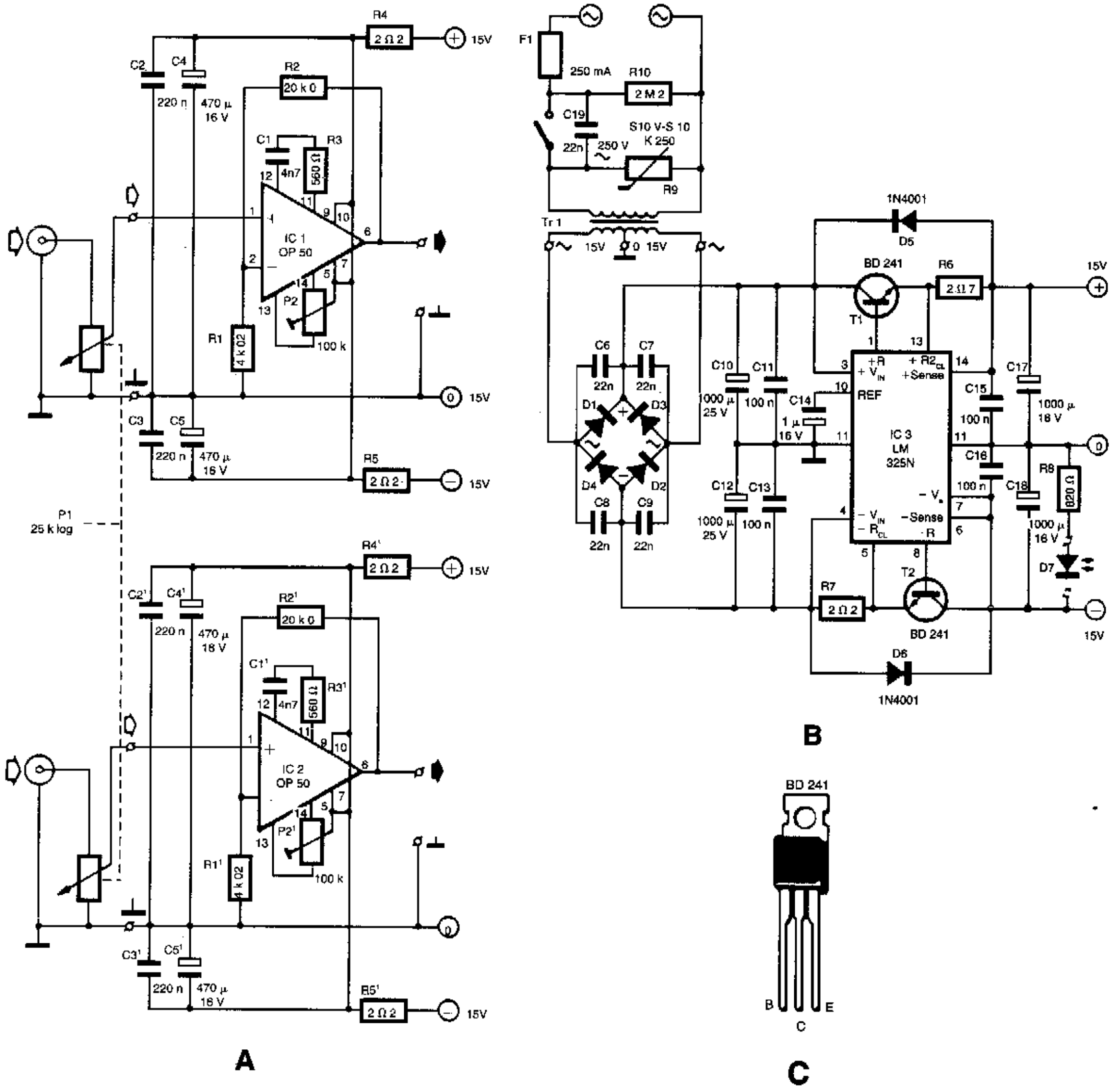
Apex Microtechnology manufactures a number of power op amps. The above circuit uses a PB50 booster amplifier to deliver a 180-V p-p signal into a 90-Ω load, from a ±100-V supply.

6

Audio Signal Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

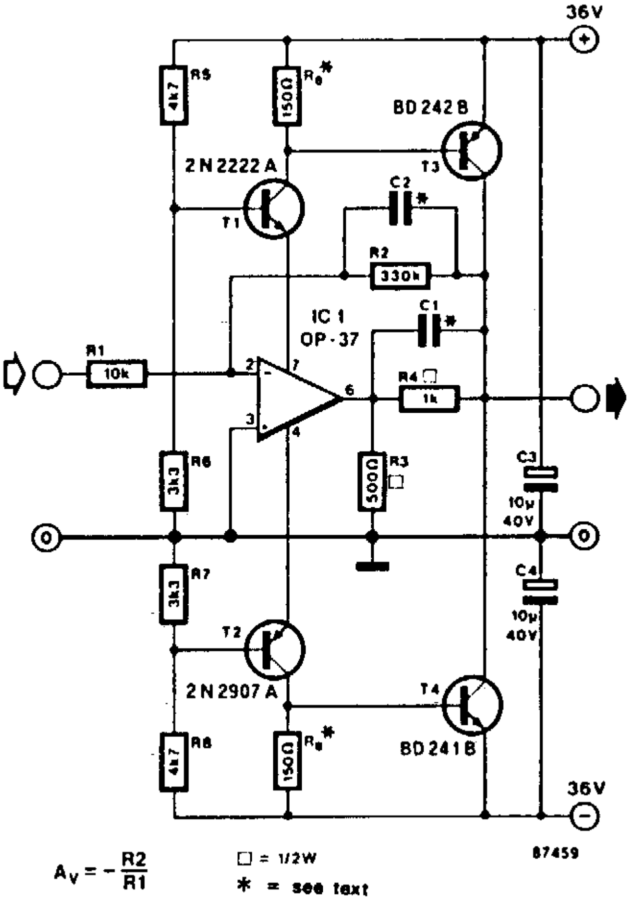
HEADPHONE AMPLIFIER



303 CIRCUITS *FIG. 6-1*

Built around Precision Monolithics Inc. OP-50 op amps, this amplifier will drive 100-Ω to 1-kΩ headphone, is flat within 0.4 dB from 10 Hz to 20 kHz, and has a THD of less than 0.01% over most of the audio range. Amplification factor is about 6X.

AUDIO LINE DRIVER

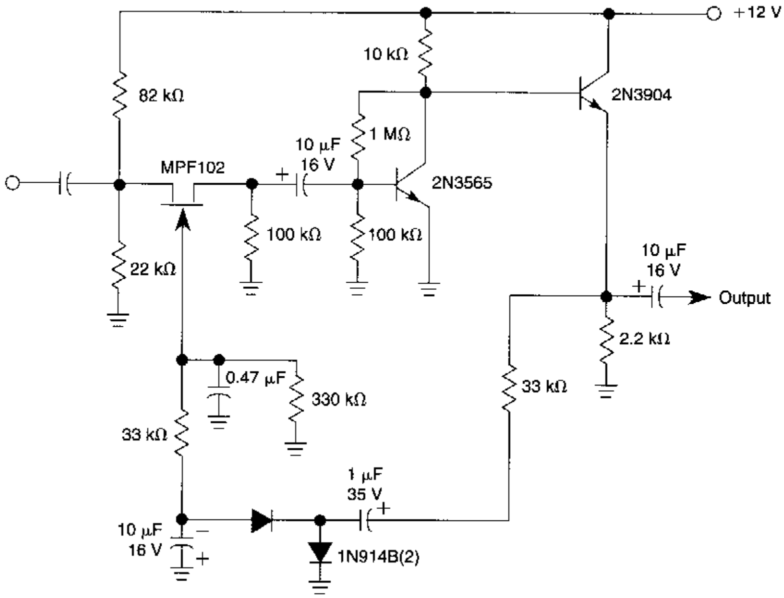


303 CIRCUITS

FIG. 6-2

This line driver can drive low-impedance lines with up to 70 V p-p max. IC1 is a low-noise op amp suitable for ±15-V operation. T1 and T2 are regulators for the power supply for IC1. T3 and T4 form a complementary power output stage. Frequency response is flat up to 100 kHz.

CONSTANT-VOLUME AMPLIFIER

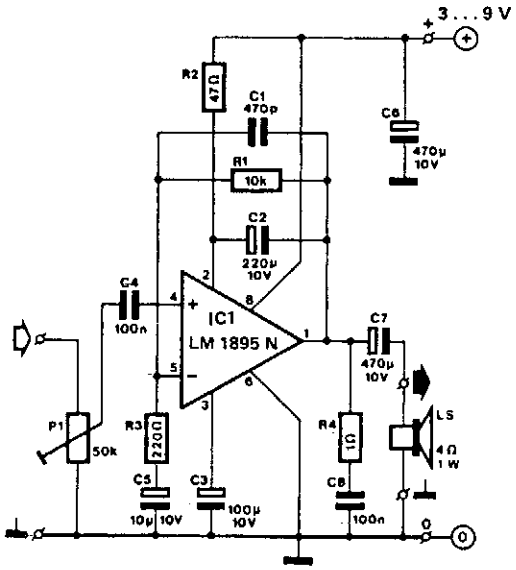


WILLIAM SHEETS

FIG. 6-3

The amplifier has an output level that shifts about 6 dB for a 40-dB input variation.

MINI AMPLIFIER USING LM1895N

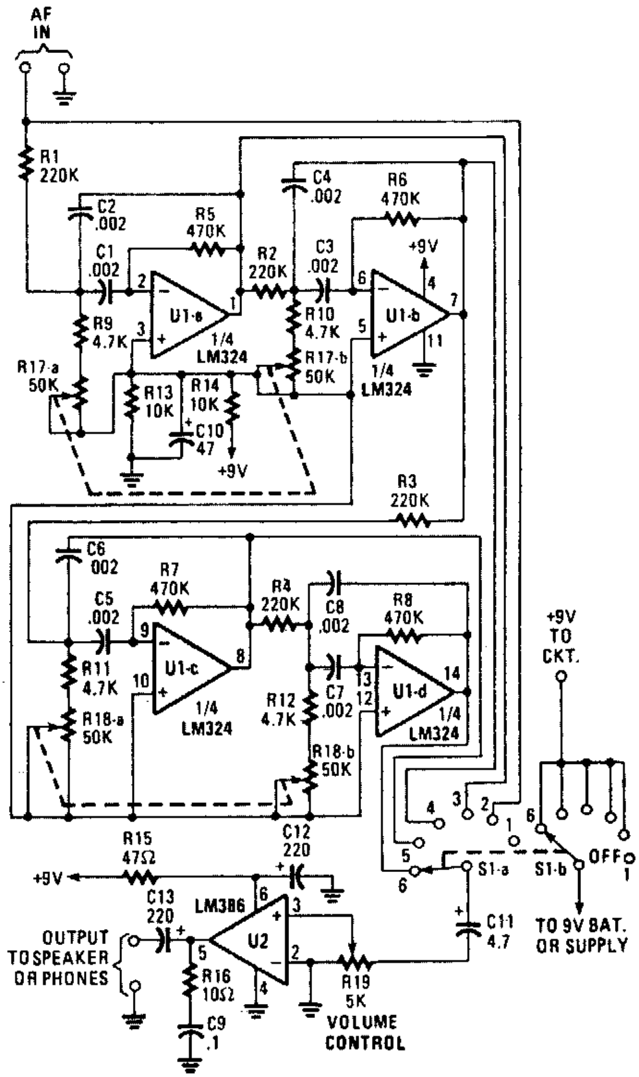


With 3-V to 9-V supplies, this amplifier can provide from 100-mW to 1-W output into a 4 Ω and bandwidth is approximately 20 kHz @ 3 dB. This circuit is useful for low-power and battery applications. Drain is 80 mA @ 3 V or 270 mA @ 9 V at maximum signal conditions.

303 CIRCUITS

FIG. 6-4

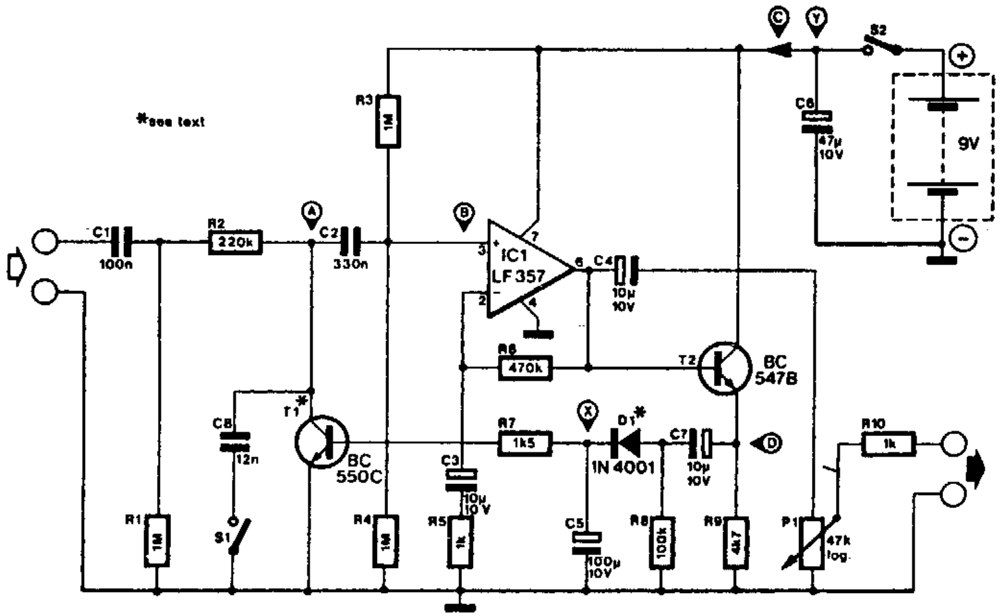
AUDIO AMPLIFIER WITH TUNEABLE FILTER



POPULAR ELECTRONICS *FIG. 6-5*

This audio amplifier can tune from 500 to 1500 Hz and will drive a speaker or headphones. Useful for CW reception or other receiver applications, only two IC devices are needed.

AUDIO COMPRESSOR

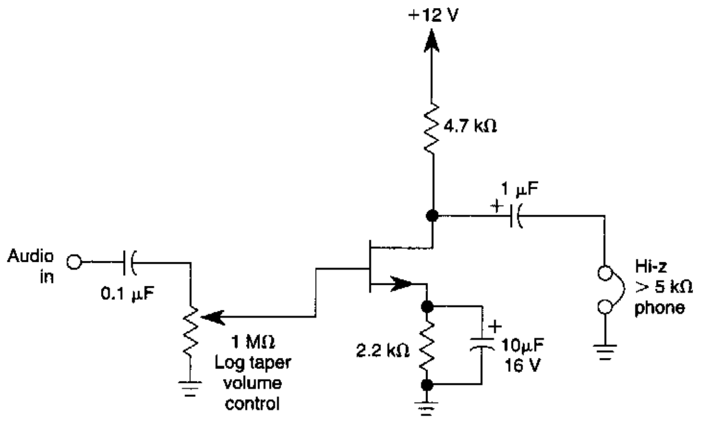


303 CIRCUITS

FIG. 6-6

This compressor will compress a 25-mV p-p to 20-V p-p audio output to input levels remaining between 1.5 V p-p to 3.5 V p-p, and has a frequency response of 7 Hz to 67 kHz. It is suitable for audio and communications applications.

JFET HEADPHONE AMPLIFIER

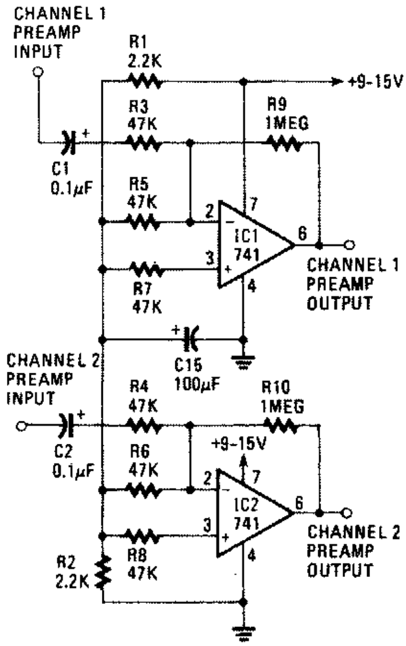


WILLIAM SHEETS

FIG. 6-7

This circuit can drive high-impedance headphones from a low impedance low-level source. Gain is about 5X to 10X depending on headphone impedance. A volume control is included.

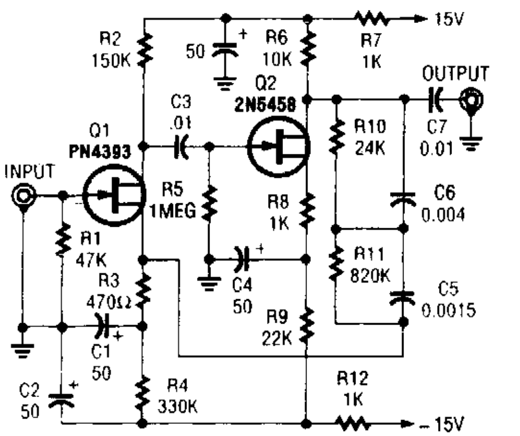
DUAL PREAMP



If you wish to amplify low-level signals, such as the output of a turntable, the signal must first be fed to this preamp.

1987 R-E EXPERIMENTERS HANDBOOK *FIG. 6-8*

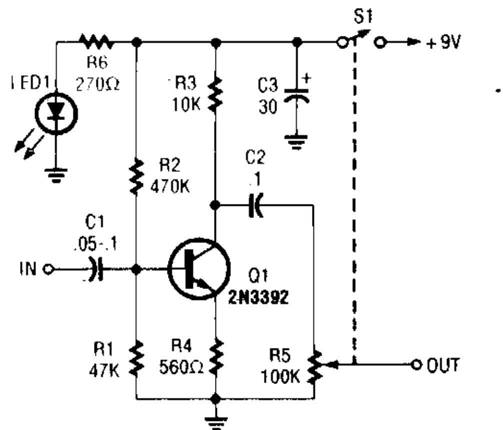
MAGNETIC PICKUP PHONO AMPLIFIER



POPULAR ELECTRONICS *FIG. 6-9*

This preamp is RAA compensated for use with magnetic phone cartridges.

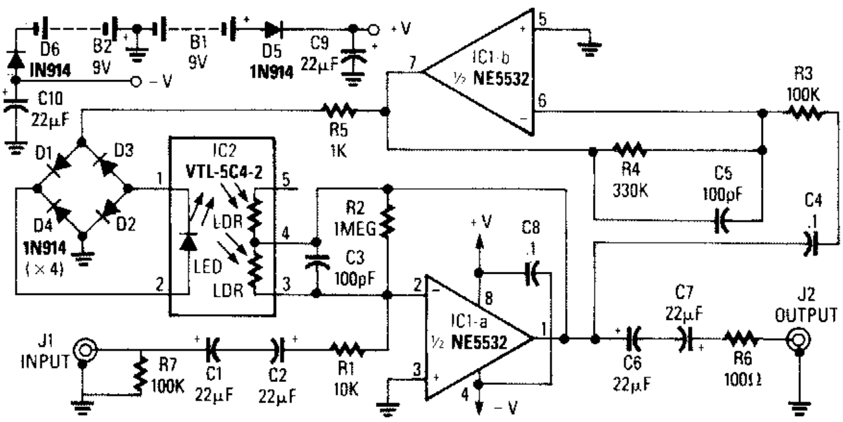
AUDIO BOOSTER



POPULAR ELECTRONICS *FIG. 6-10*

This circuit has a maximum gain of about 22 dB (voltage gain), and it can be used for miscellaneous audio circuits.

AUDIO VOLUME LIMITER

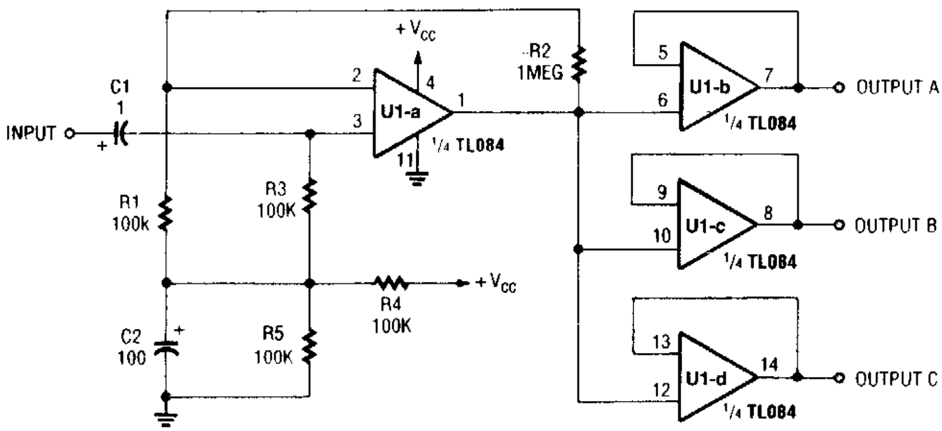


1992 R-E EXPERIMENTERS HANDBOOK

FIG. 6-11

IC1-a is connected as an inverting amplifier whose gain is controlled by the LDR portion of an optocoupler.

AUDIO DISTRIBUTION AMPLIFIER



POPULAR ELECTRONICS

FIG. 6-12

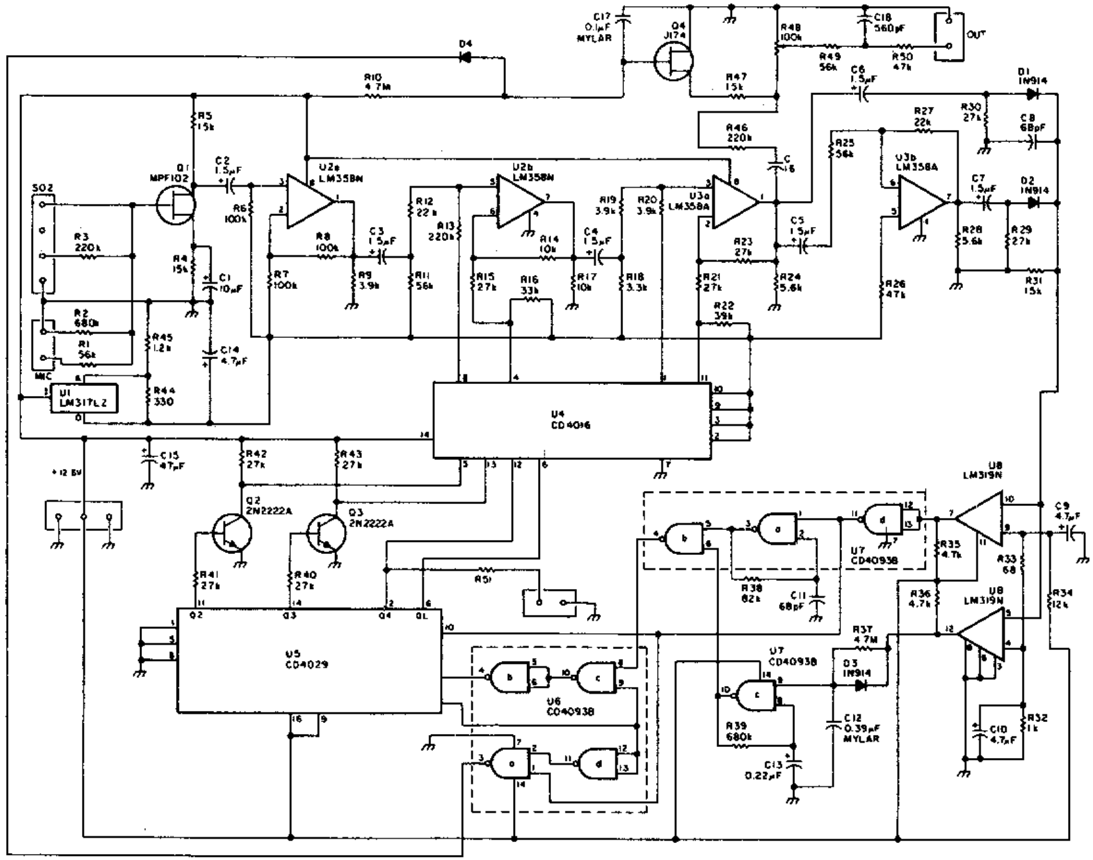
Three low-Z audio outputs are available from this circuit, using a quad TL084 FET amplifier. The input is high impedance. V_{CC} can be 6 to 12 V for typical applications.

7

Automatic Level Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

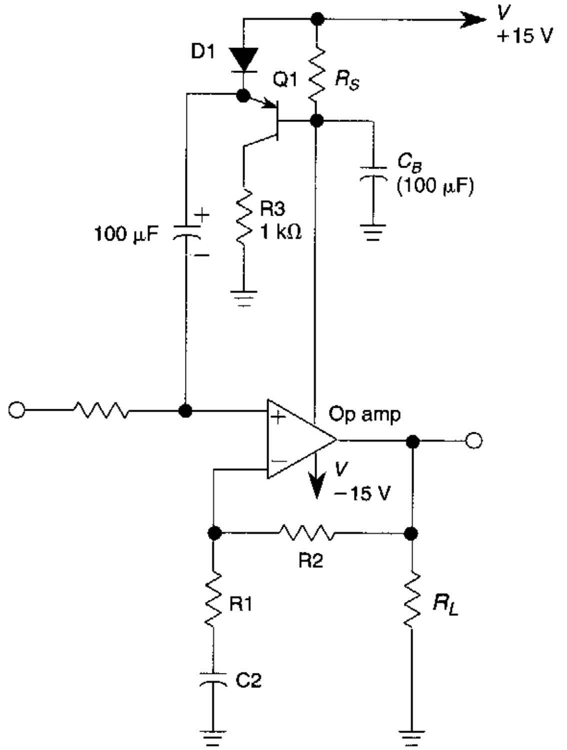
DIGITAL AUTOMATIC LEVEL CONTROL (ALC)



73 AMATEUR RADIO TODAY ***FIG. 7-1***

This approach to automatic level control (ALC) makes use of digitally switched audio attenuators in the signal path. The output level of the system is sensed, compared to a reference, and audio pads are inserted via analog switches. This method is nearly instantaneous and eliminates the compromises necessary in conventional RC network ALC systems using fast attack, slow-decay approaches.

AGC SYSTEM FOR AUDIO SIGNALS



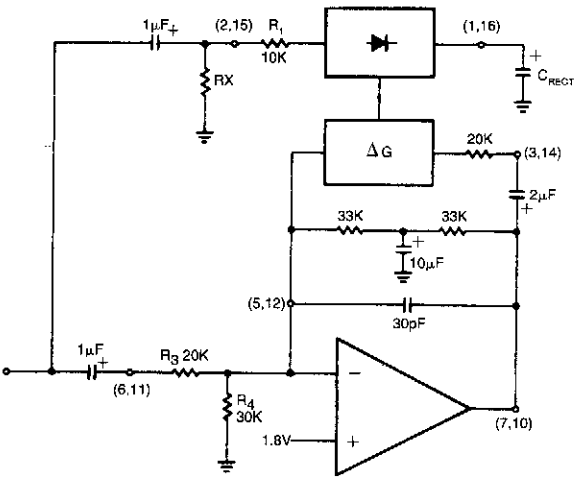
WILLIAM SHEETS

FIG. 7-2

This circuit is an AGC system for audio-frequency signals. AGC systems usually consist of three parts: an amplifier, rectifier, and controlled impedance. In this circuit the functions of an amplifier and a rectifier are performed by a single op amp. This makes the system simple and cheap.

The rectifier is made with the output push-pull cascade of the op amp and R_s, R_L, and C_B. The transistor Q1 and D1 are used as a voltage-controlled resistance (Z). The input signal is $(Z + R_1)/Z$ times, diminished by the voltage divider and $1 + R_2/R_1$ times, amplified by the op amp. C2 eliminates influence of dc bias voltage. R3 protects Q1 and D1 from excessive current.

ALC (AUTOMATIC LEVEL CONTROL)



1989 RF COMMUNICATIONS HANDBOOK

FIG. 7-3

The rectifier input is tied to the input. This makes gain inversely proportional to input level so that a 20-dB drop in input level will produce a 20-dB increase in gain. The output will remain fixed at a constant level. The circuit will maintain an output level of ±1 dB for an input range of +14 to −43 dB at 1 kHz. Additional external components will allow the output level to be adjusted.

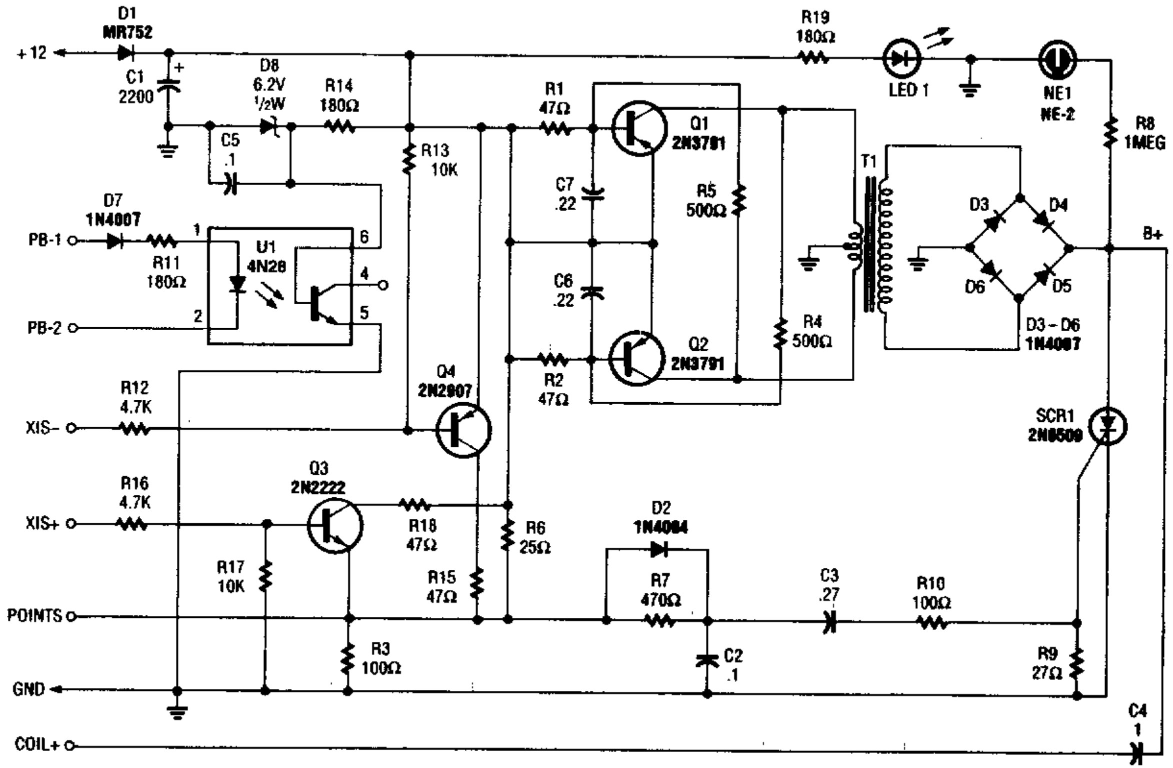
8

Automotive Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

CD Ignition System for Autos
Brake and Turn-Signal Light Circuit
Vehicular Tachometer Circuit
Smart Turn Signal
Manual Headlight/Spotlight Control for Autos
Thermostat Switch for Automotive Electric Fans
Flashing Brake Light
Power Controller (for Automotive Accessories)
Automotive Power Adapter for dc-Operated Devices
Time-Delay Auto-Kill Switch
Booster Amplifier for Car Stereo Use
Auto Turn-Signal Reminder
Headlight Flasher
Automotive Audible-Turn Indicator
Engine Block Heater Minder
Headlights-On Reminder
Brake and Turn Indicator
Lamp-Switching Circuit
Automatic Turn-Off Control for Automobiles
Alternator Regulator
Auto Generator Regulator
Lights-On Reminder
Auto Fuse Monitor
Headlight Alarm

CD IGNITION SYSTEM FOR AUTOS



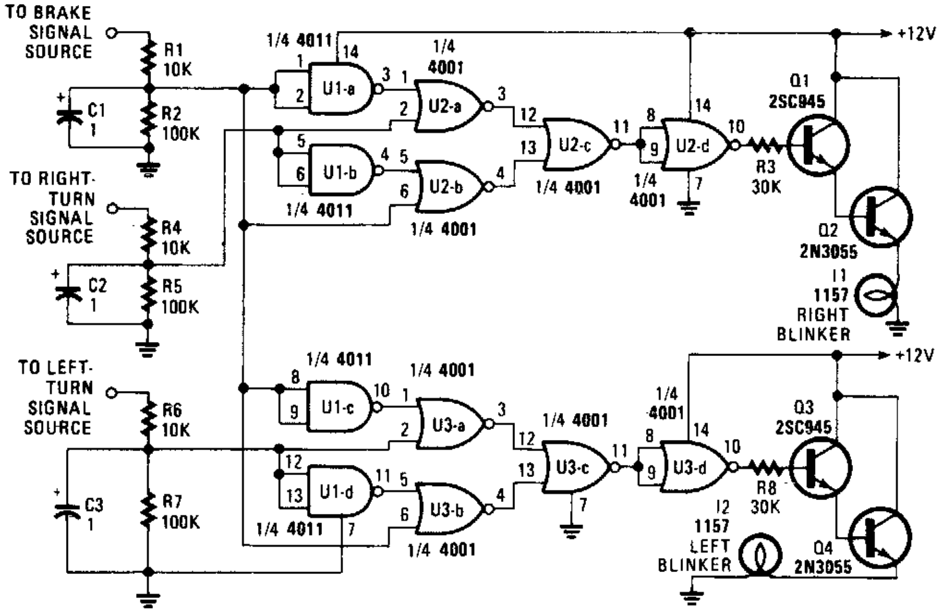
POPULAR ELECTRONICS

FIG. 8-1

At the heart of the CD4-MX is an astable multivibrator, built around Q1 and Q2, that feeds step-up transformer T1. The output of T1 is rectified by D3 to D6 and used to charge capacitor C4. When the points close, a small voltage is fed to the gate of SCR1, causing it to fire, dumping the charge of C4 to the vehicle's ignition coil. The circuit also contains optional subcircuits to accommodate different types of auto ignitions.

X_{15} + and X_{15} − are alternative trigger configurations for nonpoint breaker ignition systems. R6 is not used for these systems and must be removed. Optocoupler U1 can be used (pin 4) in conjunction with X_{15} − or X_{15} + depending on polarity of sensor. Note that 60 to 70 kV is available from this system,,so observe suitable safety precautions.

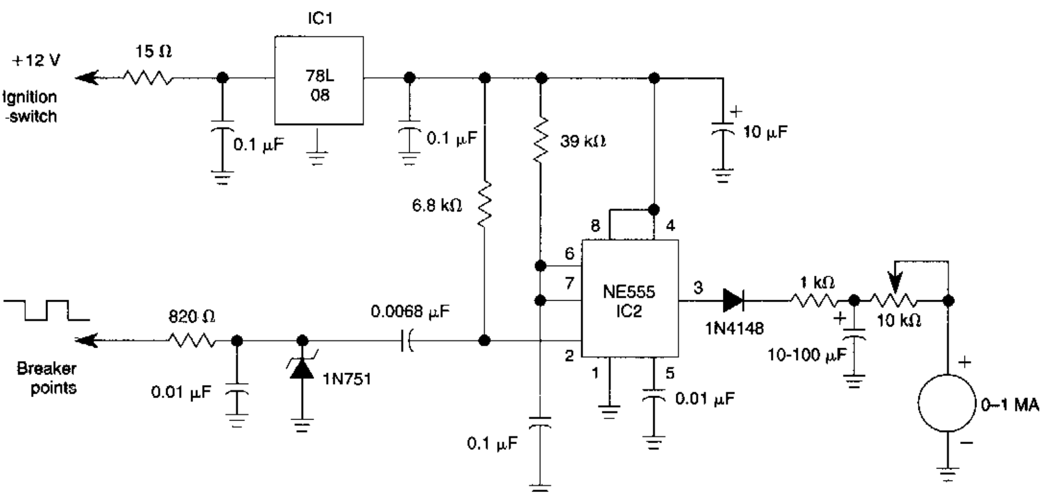
BRAKE AND TURN-SIGNAL LIGHT CIRCUIT



POPULAR ELECTRONICS ***FIG. 8-2***

This circuit enables single-filament tail-lights to serve as combination brake lights and turn signals.

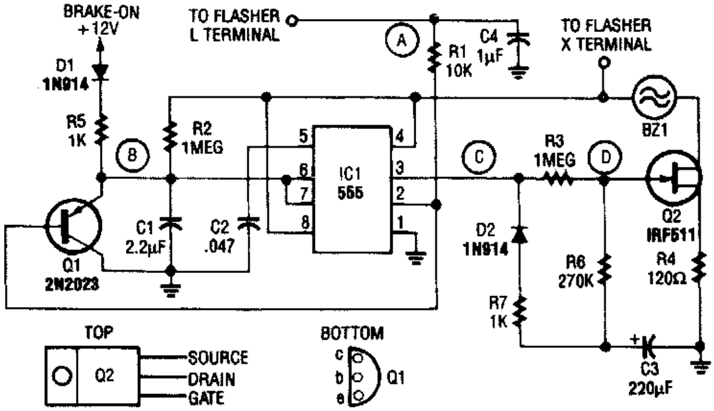
VEHICULAR TACHOMETER CIRCUIT



WILLIAM SHEETS ***FIG. 8-3***

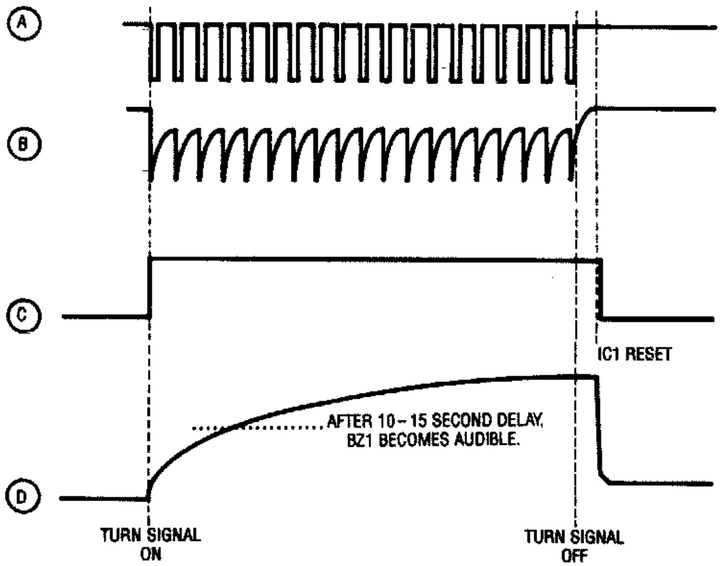
In this automotive application, the 555 is a pulse counter. IC1 regulator provides proper operating voltage for IC2. This circuit is for vehicles with conventional breaker points.

SMART TURN SIGNAL



STS schematic. The Q2 gate voltage increases with the charge on C3. After 15 seconds of charging, the buzzer will warble. As the charging continues, the sound will grow louder.

A



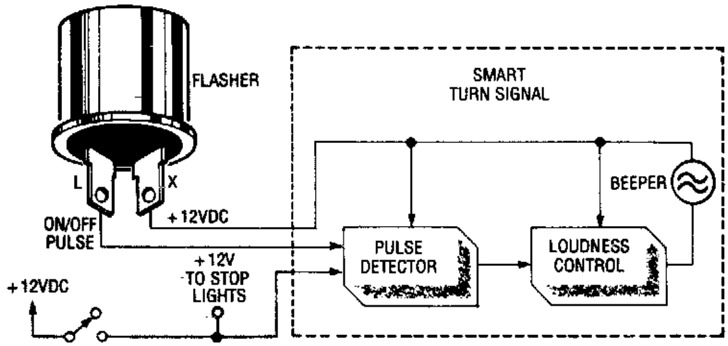
Circuit waveforms. Point A shows the signal from the flasher. The voltage at point D will increase as long as the pin-3 output of IC1 (point C) remains high. The C1-R2 time constant (point B) determines how long the output will be high.

B

ELECTRONICS NOW

FIG. 8-4

SMART TURN SIGNAL (*Cont.*)

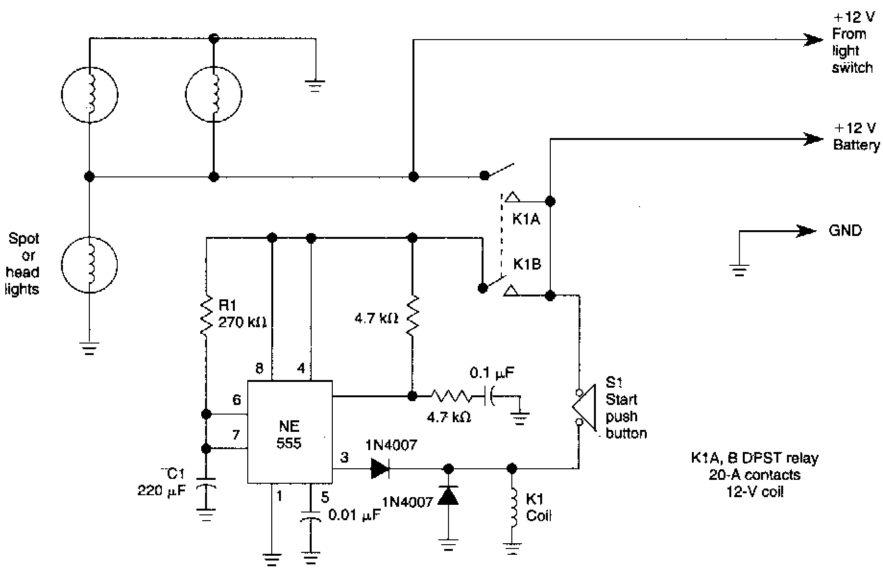


Flasher terminal L connects to the load and X connects to the 12-volt supply. When the driver engages the turn signal, the L terminal voltage varies with the blinking lights. The STS senses the changing voltage and, after 15 seconds, it applies power to a buzzer through a current-limiting device to control loudness.

C

This circuit reminds a driver that his turn signal has been left on for more than 15 seconds. When stopped for a light, the brake-on signal holds the warning off.

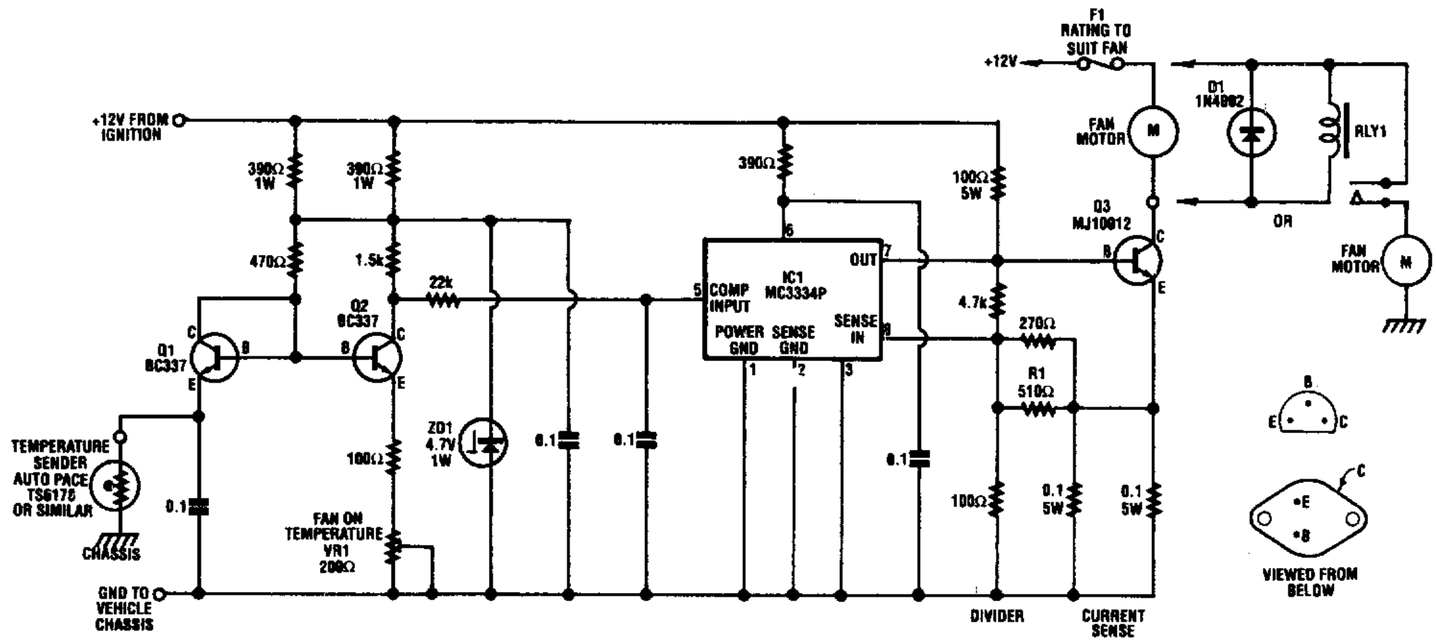
MANUAL HEADLIGHT/SPOTLIGHT CONTROL FOR AUTOS



WILLIAM SHEETS ***FIG. 8-5***

Pressing the START pushbutton turns on either the headlights or spotlights for a predetermined time. After 1 minute (R1 and C1 determine this), the lights will shut off as the NE555 completes its cycle.

THERMOSTAT SWITCH FOR AUTOMOTIVE ELECTRIC FANS

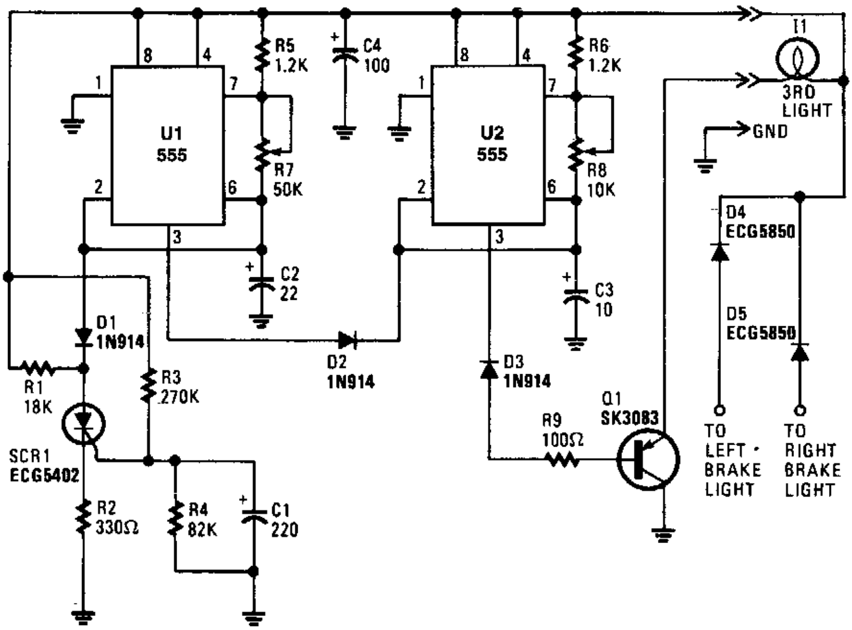


SILICON CHIP

FIG. 8-6

The circuit is based on a commercial temperature sensor (TS6178) and an MC3334P ignition chip. When the radiator temperature increases, the sensor pulls the base of Q2 low via Q1, which is wired as a diode. Q2's collector thus goes high and triggers IC1, which switches its pin 7 output high and turns on the fan motor via Q3.

FLASHING BRAKE LIGHT



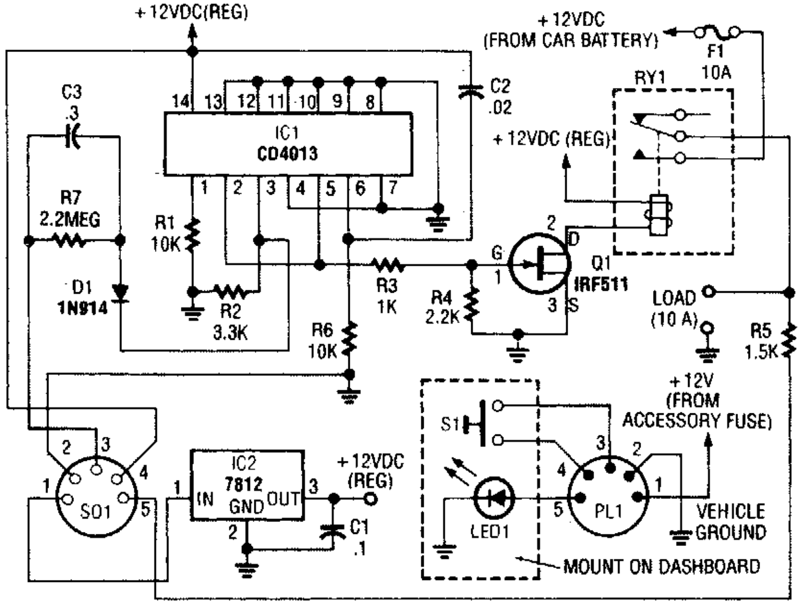
1990 PE HOBBYIST HANDBOOK ***FIG. 8-7***

When power is first applied, three things happen: the light-driving transistor (Q1) is switched on because of a low output from U2, pin 3; timer U1 begins its timing cycle, with the output (pin 3) going high, inhibiting U2's trigger (pin 2) via D2; and charge current begins to move through R3 and R4 to C1.

When U1's output goes low, the inhibiting bias on U2 pin 2 is removed, so U2 begins to oscillate, flashing the third light via Q1, at a rate determined by R8, R6, and C3. Oscillation continues until the gate-threshold voltage of SCR1 is reached, causing it to fire and pull U1's trigger (pin 2) low. With its trigger low, U1's output is forced high, disabling U2's triggering. With triggering inhibited, U2's output switches to a low state, which makes Q1 conduct, turning on I1 until the brakes are released. Removing power from the circuits resets SCR1, but the RC network consisting of R4 and C1 will not discharge immediately and will trigger SCR1 earlier. So, frequent brake use means fewer flashes.

Bear in mind that the collector/emitter voltage drop across Q1, along with the loss across the series-fed diodes, reduces the maximum available light output. If the electrical system is functioning properly (at 13 to 14 V for most vehicles), those losses will be negligible.

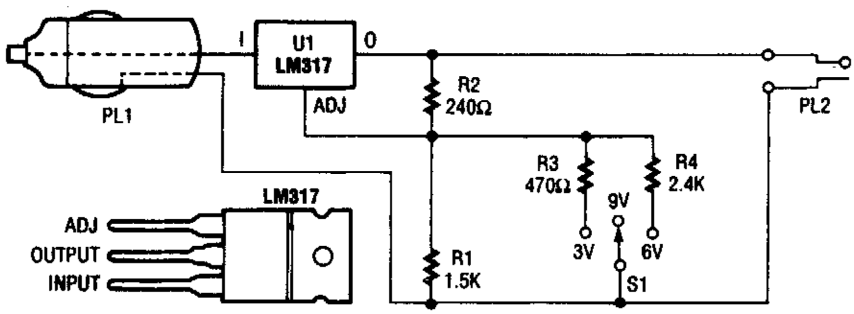
POWER CONTROLLER (FOR AUTOMOTIVE ACCESSORIES)



ELECTRONICS NOW *FIG. 8-8*

Because the power controller is powered from the vehicle's accessory switch, the load can receive power only when the ignition key is on. Using half of a dual flip-flop (CD4013), a load of up to 10 A is controlled by a momentary pushbutton. This circuit was originally intended for automotive power control, but could have other applications as well.

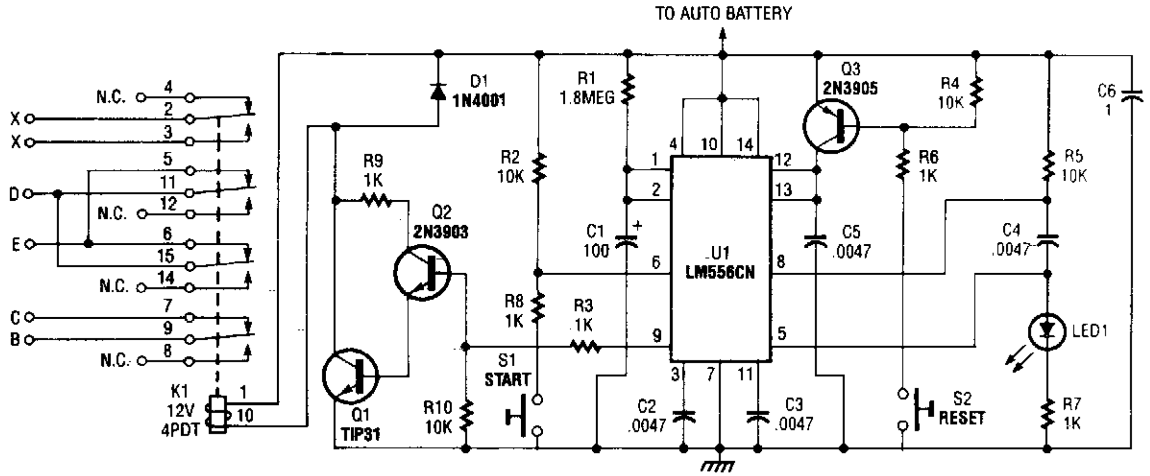
AUTOMOTIVE POWER ADAPTER FOR dc-OPERATED DEVICES



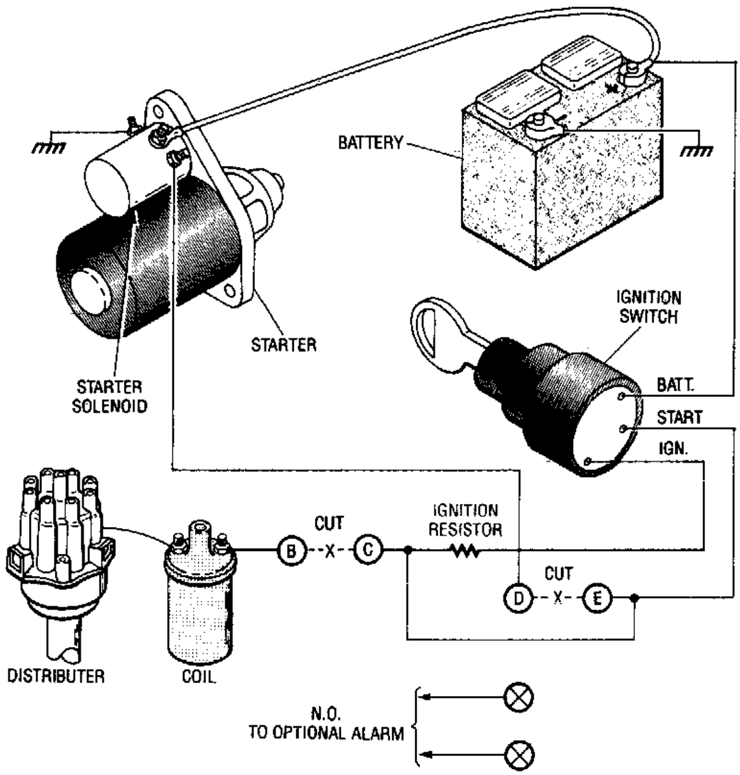
1993 ELECTRONICS HOBBYIST HANDBOOK *FIG. 8-9*

In the schematic diagram for the car-power adapter, note how the value of R_B (which is R1 and S1 in the center position) is changed by putting R3 or R4 in parallel with R1.

TIME-DELAY AUTO-KILL SWITCH



A



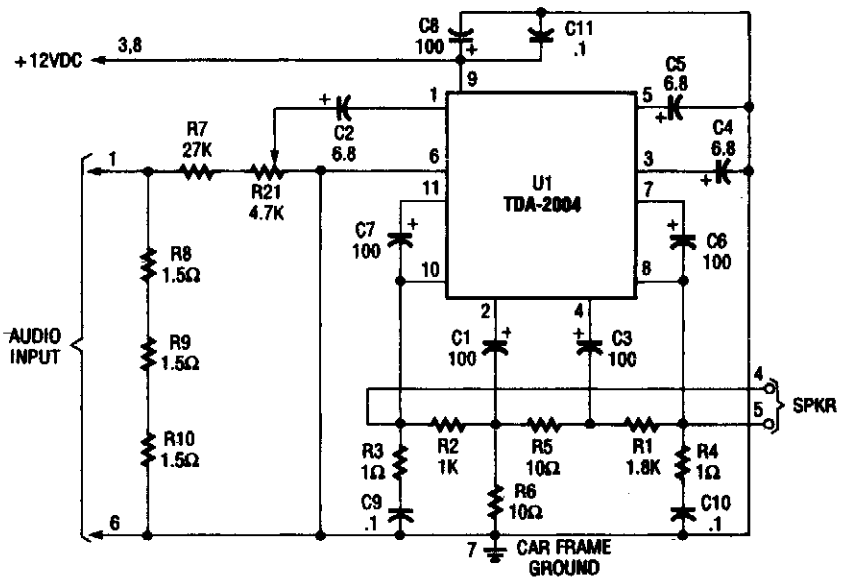
B

POPULAR ELECTRONICS

FIG. 8-10

The automobile delayed kill switch is simple in concept. When you get out of your car, a secretly located pushbutton switch is pressed. Nothing apparently happens, but at the end of a predetermined time, a relay is pulled in and locked. When the relay is pulled in, contacts open, and the hot lead from the ignition to the coil and the hot wire from the key switch to the starter solenoid is opened or disconnected. If the engine is running, it stops immediately and the starter will not operate. When you get into the car, another pushbutton switch is pressed and the relay drops out and everything goes back to normal.

BOOSTER AMPLIFIER FOR CAR STEREO USE



1990 PE HOBBYIST HANDBOOK

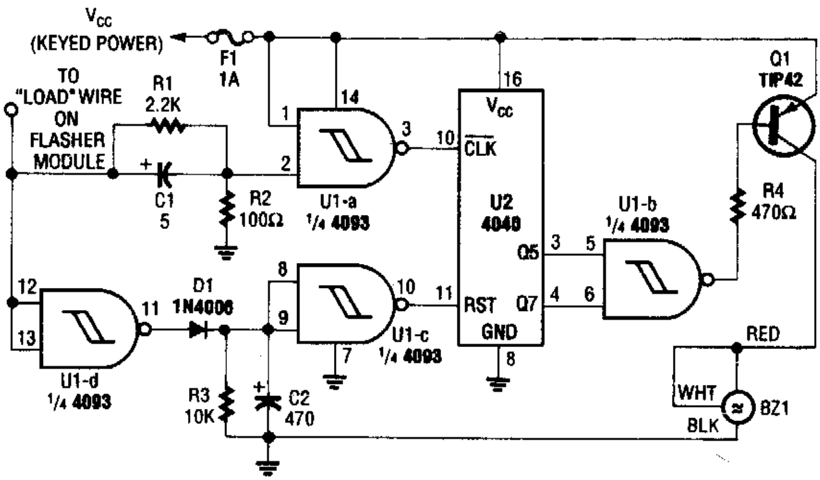
FIG. 8-11

Only one channel of this circuit is shown. The other is practically a carbon copy.

The input to the circuit, taken from your car radio's speaker output, is divided along two paths; in one path, a high-power divider network (consisting of R8 through R10) provides 4.5-Ω resistance to make the circuit's input impedance compatible with the output impedance of the car radio. In the other path, the signal is fed to the input of U1 through resistor LR7, trimmer potentiometer R21, and capacitor C2. Together, R7 and R21 offer a minimum resistance of 27,000 Ω.

Integrated circuit U1 (a TDA-2004 audio power amplifier) amplifies the signal, which is then output at pins 8 and 10 and fed to the loudspeaker. Note: This amp is designed for use only with car radios whose speaker outputs are referenced to ground: do not use it with radios that have balanced outputs.

AUTO TURN-SIGNAL REMINDER

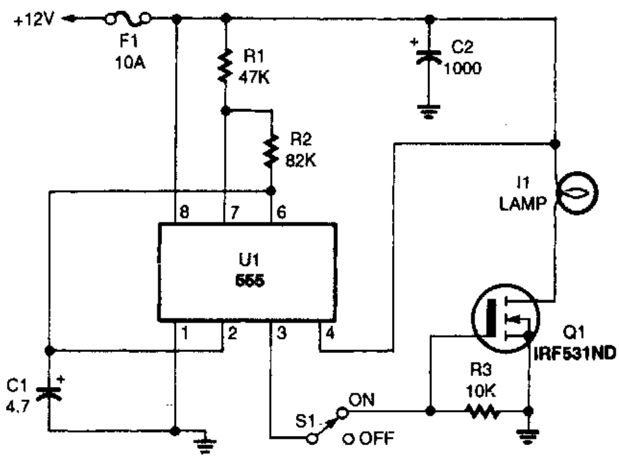


POPULAR ELECTRONICS

FIG. 8-12

This circuit counts turn signal flashes. At the end of about 70 flashes, a chime sounds to remind the driver to turn off the turn signal. By using various taps on U2, the period can be changed if desired. BZ1 is a buzzer or chime module.

HEADLIGHT FLASHER

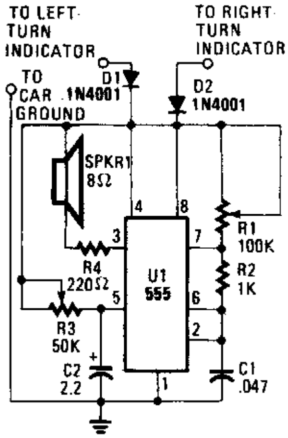


POPULAR ELECTRONICS

FIG. 8-13

The headlight flasher is nothing more than a 555 oscillator/timer that's configured as an astable multivibrator (oscillator). Its input is used to drive the gate of an IRF53IND hexFET, which, in turn, acts like an on/off switch, turning the lamp on and off at the oscillating frequency (1 Hz).

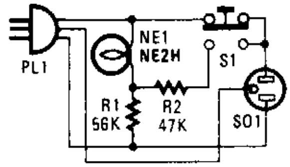
AUTOMOTIVE AUDIBLE-TURN INDICATOR



POPULAR ELECTRONICS *FIG. 8-14*

This little circuit should be useful to the hearing impaired. It produces a tone each time a dashboard turn indicator lights. The tone drops in frequency for as long as the indicator is lit.

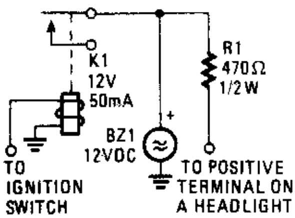
ENGINE BLOCK HEATER MINDER



POPULAR ELECTRONICS *FIG. 8-15*

If you live in the frozen north, knowing your engine-block heater is working is a comfort. This device will let you know if yours is okay. Plug in PL1 to your power outlet. NE1 should light. Then, plug in the block heater. Depressing S1 should cause the indicator to get brighter. If not, your block heater might be open and inoperative.

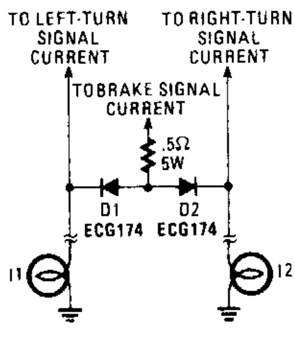
HEADLIGHTS-ON REMINDER



POPULAR ELECTRONICS *FIG. 8-16*

This circuit will sound alarm BZ1 if the ignition is turned off with the headlights on.

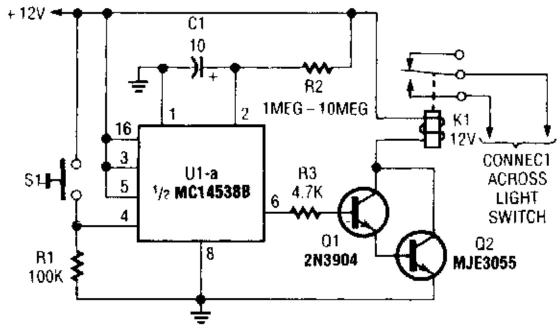
BRAKE AND TURN INDICATOR



POPULAR ELECTRONICS *FIG. 8-17*

This might be a quick solution to getting the two-wire truck harness to support both turn and braking indications.

LAMP-SWITCHING CIRCUIT

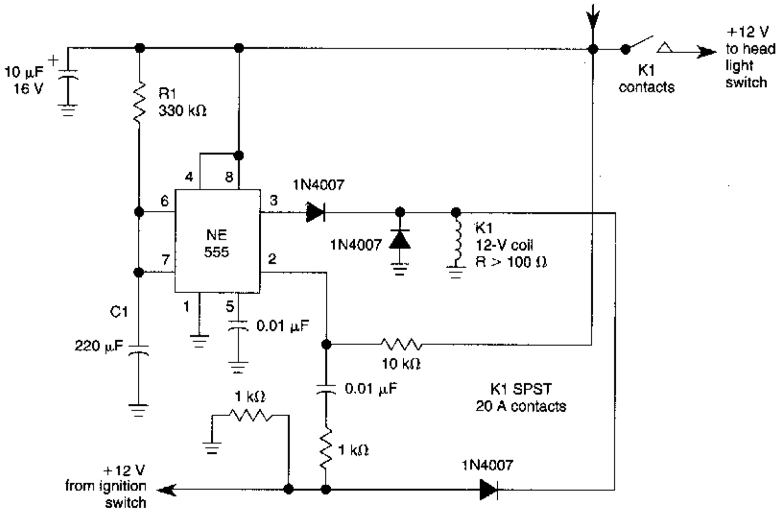


POPULAR ELECTRONICS **_FIG. 8-18_**

A normally open pushbutton switch (S1) delivers a positive input pulse to pin 4 of U1, triggering the IC into action. The output of U1 at pin 6 supplies base-drive current to a Darlington pair comprised of Q1 and Q2, activating K1. A 10-µF capacitor and any resistor value of from 1 to 10 MΩ can be used as the timing components.

To use the circuit on an auto's headlights, connect the relay's normally open contacts across the car's headlight switch and press S1 to extend the on time. In connecting the circuit to control an ac-operated lamp, turn off the ac power and connect the relay contacts in parallel with the lamp's power switch contacts.

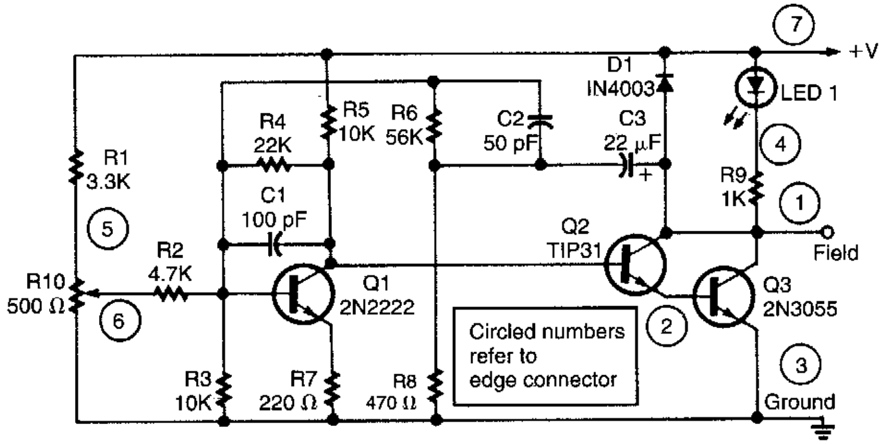
AUTOMATIC TURN-OFF CONTROL FOR AUTOMOBILES



WILLIAM SHEETS **_FIG. 8-19_**

When the ignition switch is on, relay K1 is energized continuously, and the headlights can be turned on. Turning off the ignition turns on timer IC1,which keeps IC1 energized for a time determined by R1 and C1. With the values shown approximately a 1 minute delay will result. The values of R1 or C1 can be changed to vary this delay time.

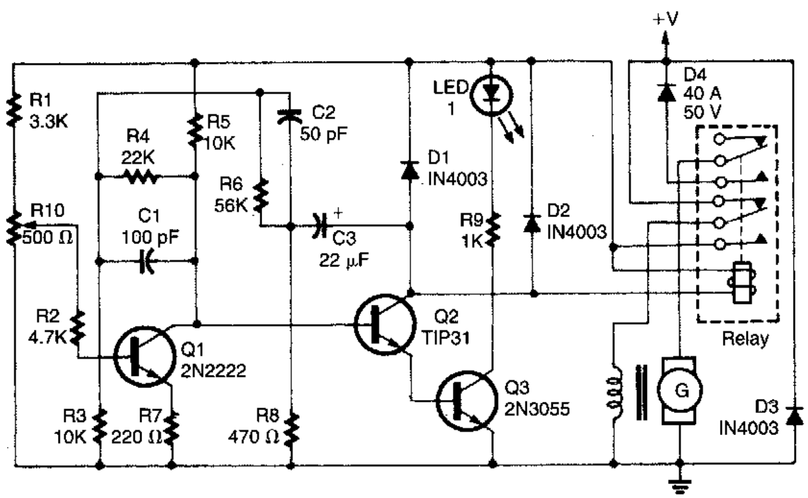
ALTERNATOR REGULATOR



RADIO-ELECTRONICS ***FIG. 8-20***

This alternator regulator uses a 3-transistor dc amplifier, and is designed for a "pulled up" field system, where one side of the alternate field returns to the +12-Vsupply, and the other end is pulled toward ground. The circuit monitors the state of the battery through a resistive divider and causes the voltage to change at the field terminal.

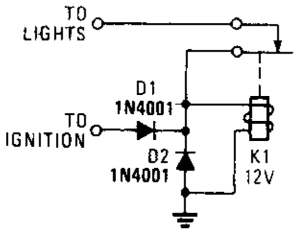
AUTO GENERATOR REGULATOR



RADIO-ELECTRONICS ***FIG. 8-21***

This regulator is for the purpose of controlling a dc generator. The field configuration is that one side of the field is grounded. D4 prevents the battery from discharging through the generator and takes the place of the mechanical cut-out relay. R10 adjusts the system voltage setting.

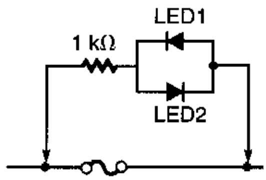
LIGHTS-ON REMINDER



POPULAR ELECTRONICS *FIG. 8-22*

A relay and two diodes are all that is needed—the relay performs the job of a buzzer so no annunciator is required. When the lights are left on, but the ignition is off, the normally closed relay contacts are in series with the relay coil. That means the relay interrupts its own power each time it becomes active, so it chatters and acts like a buzzer. This is a real minimalistic headlight reminder. It doesn't even require an annunciator because the relay acts as buzzer.

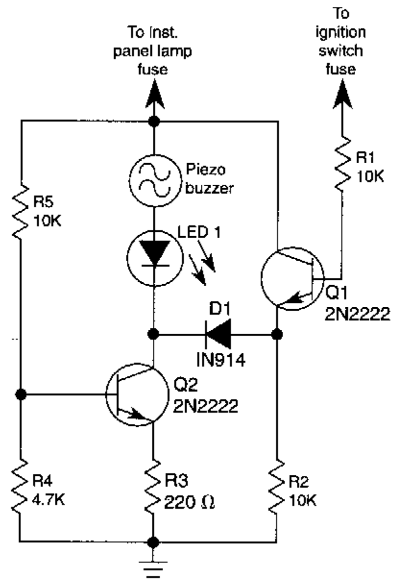
AUTO FUSE MONITOR



WILLIAM SHEETS *FIG. 8-23*

This circuit can quickly check a fuse in an automobile circuit. Connect across suspected fuse—either LED glows, fuse is blown. The circuit must be live for this test to work.

HEADLIGHT ALARM



1989 R-E EXPERIMENTERS HANDBOOK *FIG. 8-24*

The base of Q1 is connected to the car's ignition circuit; the easiest point to make that connection is at the ignition switch fuse in the car's fuse panel. Also, one side of the piezoelectric buzzer is connected to the instrument-panel light fuse; when the headlights or parking lights are on, the instrument panel is lit, too. When the headlights are off, no current reaches the buzzer. Therefore, nothing happens. What happens when the headlights are on depends on the state of the ignition switch. When the ignition switch is on, transistors Q1 and Q2 are biased on, effectively removing the buzzer and the LED from the circuit.

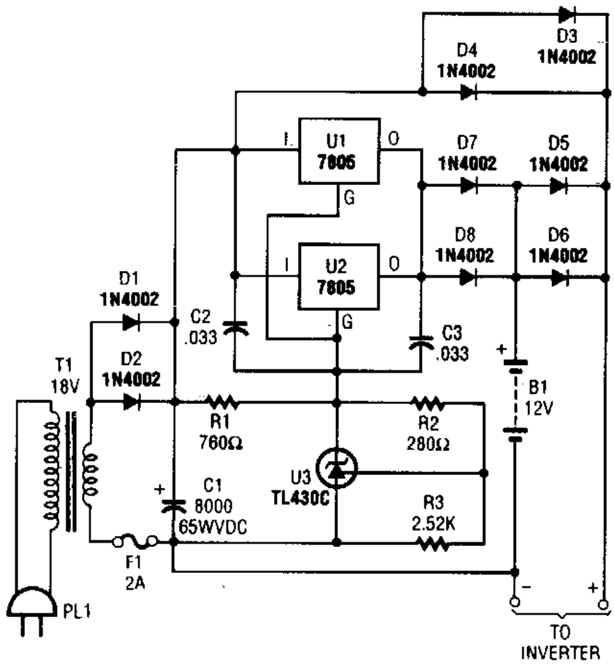
When the ignition switch is turned off, but the headlight switch remains on, transistor Q1 is turned off, but transistor Q2 continues to be biased on. The result is that the voltage across the piezoelectric buzzer and the LED is sufficient to cause the buzzer to sound loudly and the LED to light.

9

Battery Charger Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

LEAD-ACID TRICKLE CHARGER



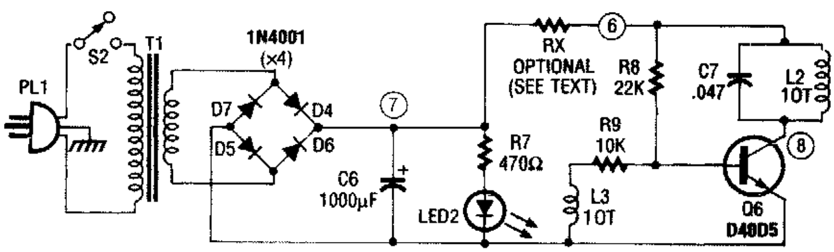
POPULAR ELECTRONICS

FIG. 9-1

The charger can be used as a stand-alone charger or for emergency lighting and burglar alarm systems using lead-acid batteries.

RF-TYPE BATTERY CHARGER

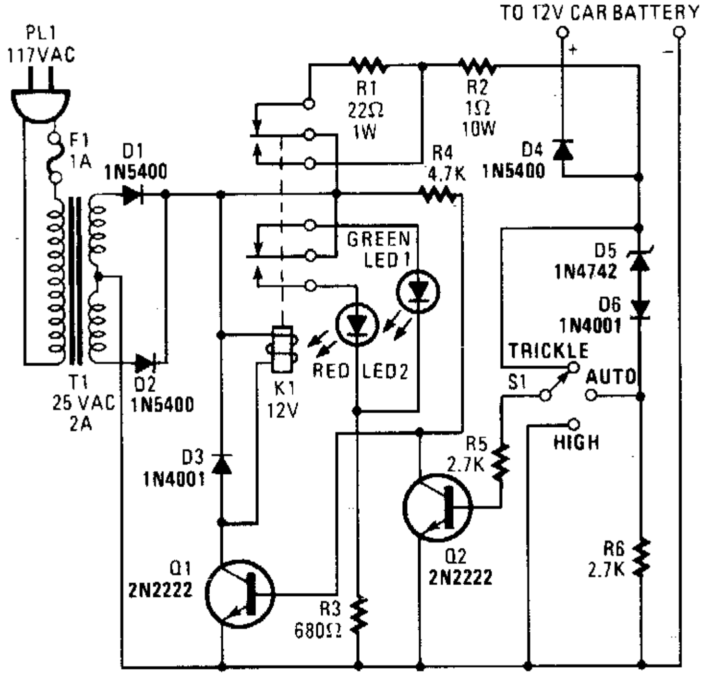
This type of charger couples RF from L2 to an external pickup coil. The pickup coil connects to a rectifier and battery to be charged. This idea is handy because no wire or contacts are required. L2 is 10T #24 wire and L3 is 10T #30 wire. Both coils are mounted on a 1" × ¼" ferrite rod.



1992 R-E EXPERIMENTERS HANDBOOK

FIG. 9-2

BATTERY CHARGER



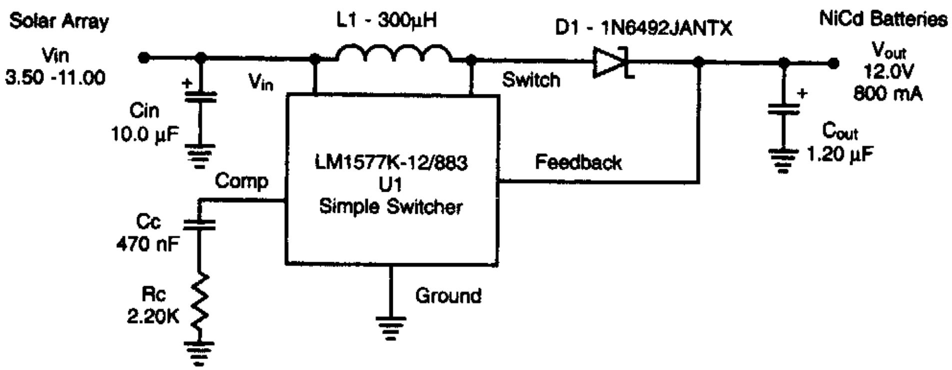
POPULAR ELECTRONICS *FIG. 9-3*

The circuit is capable of supplying either a trickle (50 mA) or high-current (1-A) charge. You can select either charging method or an automatic mode that will first trickle charge a battery if it is particularly low before switching to high-current charging.

If the battery's voltage is low, Zener-diode D5 will not conduct sufficient current to produce a voltage drop across R6 to turn Q2 on. With Q2 off, R4 pulls the base of Q1 high, turning it on. That activates K1. With K1 active, the only thing between the battery and the power supply is R2 and D4 (which prevents current from flowing through the circuit from the battery).

Once the battery charges a bit, the current through D5 increases, causing a voltage drop across R6 that is of sufficient magnitude to turn on Q2. Transistor Q2, in turn, grounds the base of Q1, keeping it off. With Q1 off, K1 remains in its normally closed state. That places R1 in series with the battery, thereby reducing the current to a trickle.

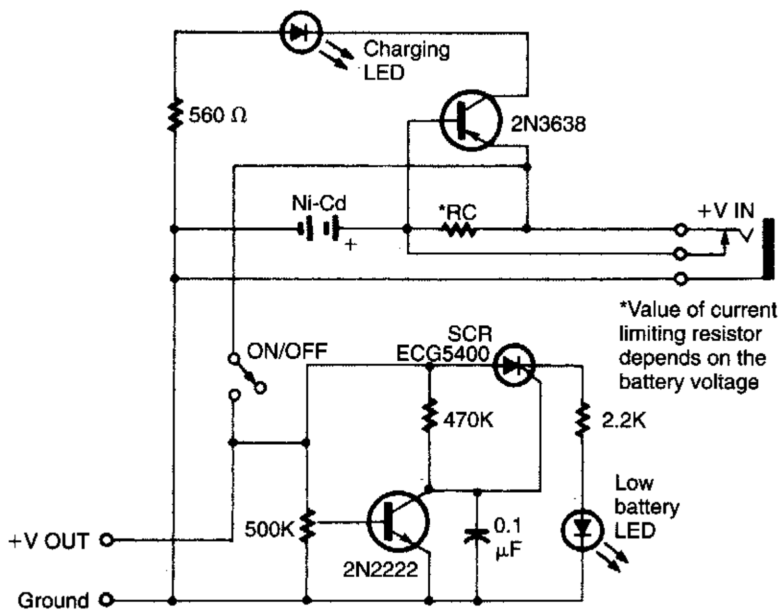
SOLAR-POWERED BATTERY CHARGER



NATIONAL SEMICONDUCTOR *FIG. 9-4*

A National Semiconductor LM1577 IC is used in a step-up regulator to charge Nicad batteries from a solar panel.

INTELLIGENT BATTERY-CHARGING CIRCUIT



RADIO-ELECTRONICS *FIG. 9-5*

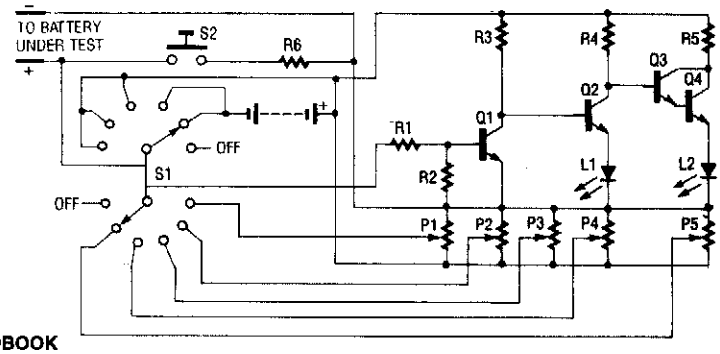
Intended for a Nicad application this charging circuit can be used with a wide range of batteries. A low-battery detector is intended. The trip voltage is set via the 500-kΩ pot. Select R_C for the battery you intend to use.

10

Battery Test and Monitor Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

BATTERY TESTER



1991 PE HOBBYIST HANDBOOK

FIG. 10-1

The battery tester uses four transistors and two LEDs to indicate the condition of any battery you want to test. Q3 and Q4 are connected in a Darlington configuration that has extremely high gain. LED L2 lights when a small positive potential appears on the base of Q3. Transistors Q1 and Q2 form a direct-coupled dc-amplifier circuit. The output of this stage drives the red LED L1. Rotary switch S1 is used to select different ranges (which have been previously set by adjusting trimmer resistors P1 through P5).

The positive (+) lead goes through the selected contacts of S1 to the biasing resistors R3, R4, and R5. The negative (–) lead of the battery under test goes to the ground or common lead of the circuit and the (+) side to one side of P1 through P5.

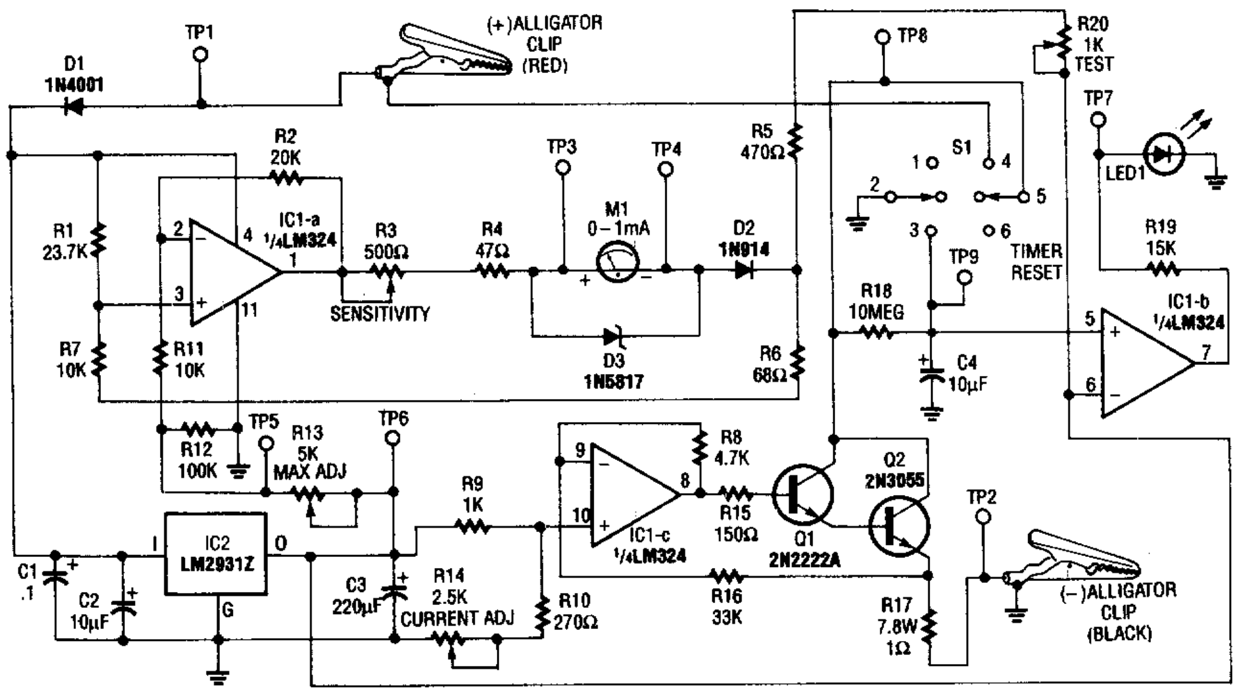
L1	Red LED
L2	Green LED
P1 through P5	5-kΩ trimmer resistor
R1	100 kΩ
R2, R3	33 kΩ
R4, R5	470 Ω
R6	12 Ω 1 W
S1	2 P6 position NS rotary switch
S2	NO pushbutton switch

Depending on the position of S1, a particular trimmer resistor (wiper lead) is selected. That lead goes through the contact on S1 to resistor R1 and into the base of npn transistor Q1. If the battery is good enough, (+) voltage goes to the base of Q1, turning it on. This turns Q2 off, which then allows Q3 to turn on. That causes Q4 to turn on and light green LED L2.

If the battery is weak, Q1 will not turn on, which will cause Q2 to be biased on by R3, which in turn lights red LED L1. When Q1 is on, it biases the base of Q3 negative, and causes Q3 to be turned off. That prevents L2 from turning on.

The circuit operates in the same manner for all ranges except the first two, where a 9-V battery has been added by S1 to be in series with the input voltage to allow for testing of very low voltage batteries. That is because at voltages below 2 Vdc, LEDs will not light and the circuit would be unable to set a low-voltage (<2-V) battery without the additional internal-battery voltage. A load resistor has also been included; it allows the battery under test to be connected to a load to give a better indication of its condition. That load resistor is connected across the battery when normally open (NO) switch S2 is depressed.

CAR BATTERY TESTER FOR CRANKING AMPS

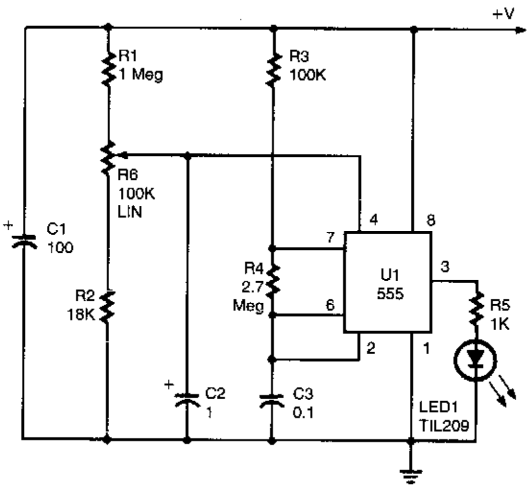


ELECTRONICS NOW

FIG. 10-2

This circuit determines the cold cranking amps of a battery by first discharging the surface charge, then checking the internal resistance. This gives a more realistic measurement than simply measuring the instantaneous drop in voltage with a load. A constant-current source draws 2.5 A. Then, after one minute, a voltage drop measurement is made under load.

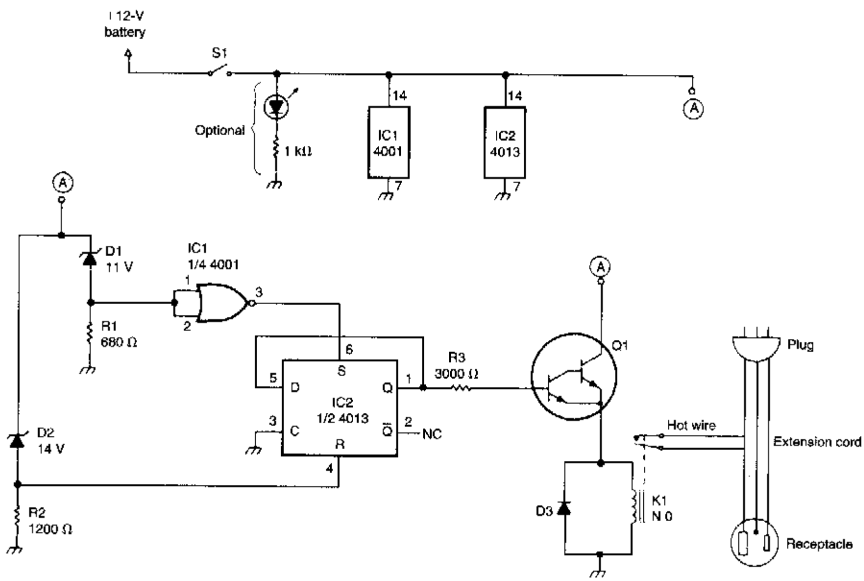
SUPPLY VOLTAGE MONITOR



When supply voltage exceeds a preset level, the 555 oscillates, and flashes LED1. The flash rate is controlled by varying C3.

POPULAR ELECTRONICS *FIG. 10-3*

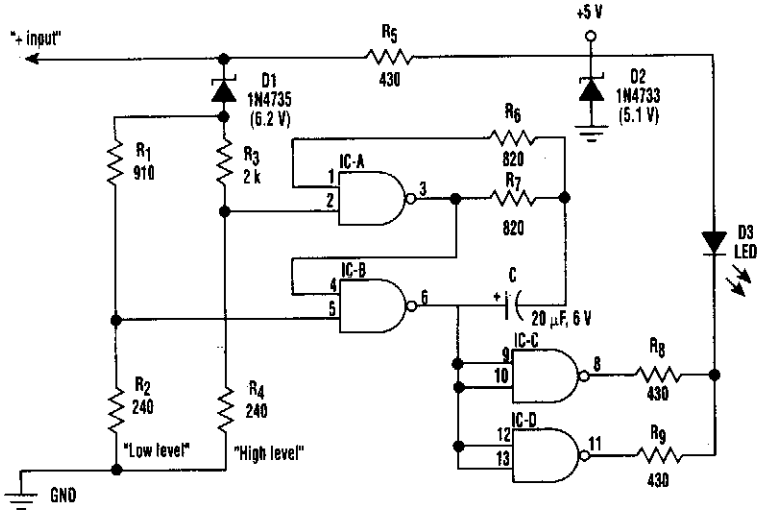
BATTERY WATCHDOG



73 AMATEUR RADIO TODAY *FIG. 10-4*

This circuit uses a pair of Zener diodes to monitor battery voltage of a 12-V battery. If below 11 V, D1 ceases to conduct, pin 3 of IC2 goes high, setting FF IC2 turning on Q1, K1, and the battery charger. At excess of 14-V battery voltage (full charge), D2 conducts, resetting FF IC2, and cutting off the battery charger.

BATTERY TEST CIRCUIT

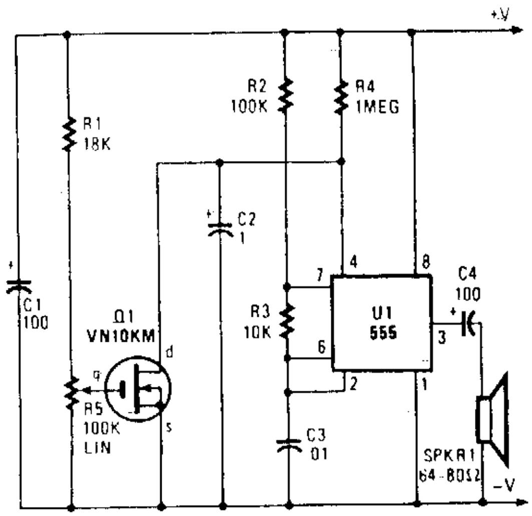


ELECTRONIC DESIGN

FIG. 10-5

Using this circuit, three levels of voltage can be displayed—normal (11 to 15 V), high (>15 V), and low (<11 V). When the voltage is low, the LED glows steadily. In the normal range, the LED is off. When the voltage is high, the LED blinks at a 1-Hz rate. This circuit is useful for assuring proper electrical system operation.

BATTERY VOLTAGE MONITOR

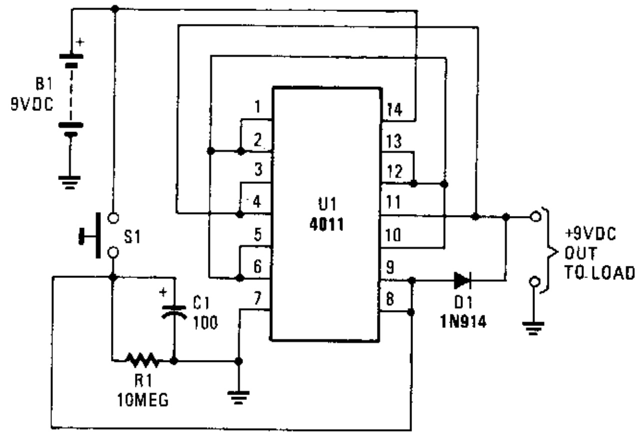


When battery voltage goes low, pin 4 of U1 goes high as Q1 fails to conduct. This activates oscillator U1 and generates audio tone. R5 sets level at which the circuit activates.

POPULAR ELECTRONICS

FIG. 10-6

BATTERY SAVER CIRCUIT

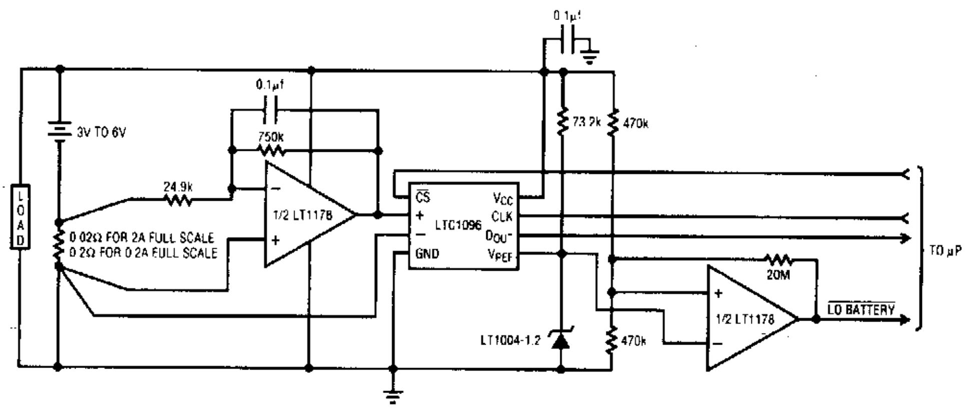


POPULAR ELECTRONICS *FIG. 10-7*

This battery saver circuit can automatically turn off a small piece of test equipment after a desired period of time, allowing you to leave your shop worry free.

This circuit uses a CD4011 IC to act as a simple timer. One section acts as an RC discharge timer (pin 7). This causes its output to go low, holding the three other outputs high acting as a 9-V source. After C1/R1 discharges approximately 10 minutes, the output drops to zero. S1 resets the circuit.

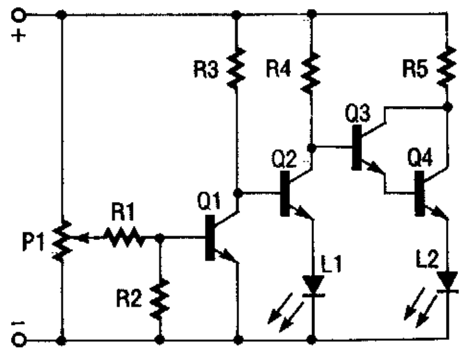
0–2-A BATTERY CURRENT MONITOR WITH DIGITAL OUTPUT



LINEAR TECHNOLOGY *FIG. 10-8*

IC devices by Linear Technology make up this current monitor circuit. Drain is only 70µA from a 3- to 6-V battery.

CAR BATTERY AND ALTERNATOR MONITOR



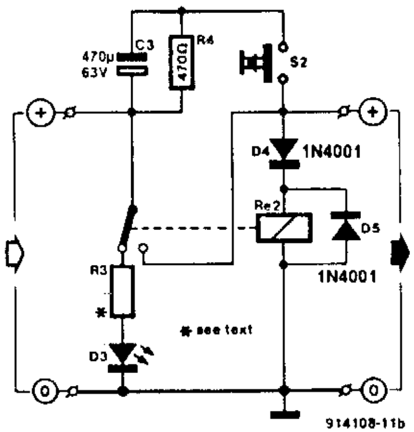
L1	Red LED
L2	Green LED
P1	2.5-kΩ trimmer resistor
Q1–Q4	2N3904 transistor
R1	100-kΩ resistor
R2, R3	33-kΩ resistor
R4, R5	470-Ω resistor
Misc.	PC board, wire

1991 PE HOBBYIST HANDBOOK ***FIG. 10-9***

The monitor is a simple voltage comparator in which a car battery serves as the battery for operation. The input voltage to the comparator is set by adjustment potentiometer P1, which must be adjusted so that the green LED L2 is on when the alternator is operating properly and red LED1 is on when the alternator is inoperative.

The circuit operates as follows: When the alternator operates properly, the battery voltage is higher and P1 is set so that transistor Q1 causes Q2 to be off. That results in Q3 and Q4 being fully on, thus applying current to green LED L2. If the battery voltage is lowered (alternator inoperative), transistor Q1 is turned off. That allows transistor Q2 to turn fully on, applying current to red LED L1, indicating trouble. Once Q2 is on, it causes Q3 and Q4 to go out of conduction.

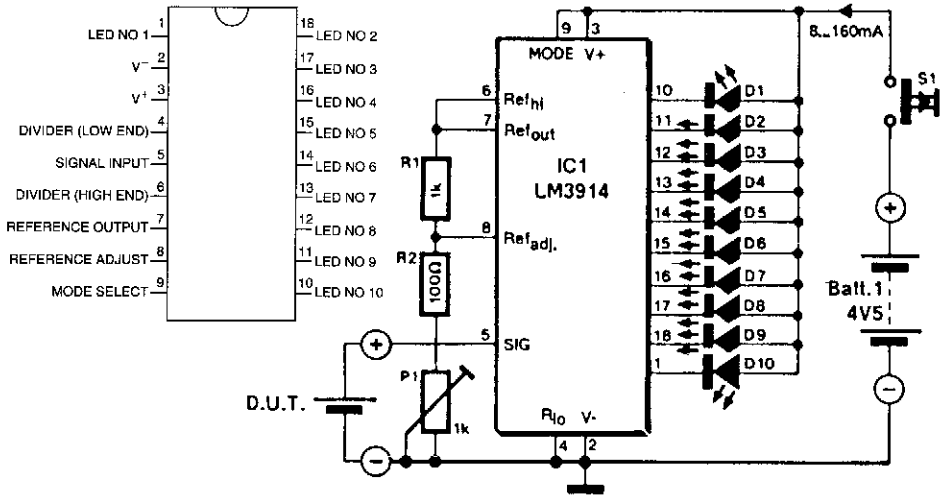
RELAY FUSE FOR BATTERY CHARGES



ELEKTOR ELECTRONICS ***FIG. 10-10***

Charged capacitor C3 and momentary push-button switch S2 are used to momentarily energize relay RE2. The battery under charge energizes the relay to hold it closed. S2 will energize the relay even if the battery is too far discharged initially to energize it.

BARGRAPH LED BATTERY TESTER



ELEKTOR ELECTRONICS USA *FIG. 10-11*

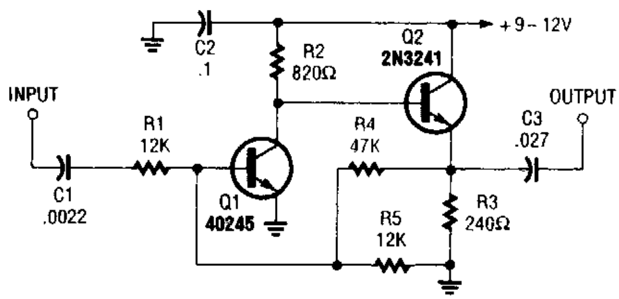
The LM3914A bargraph LED is used here as a voltmeter for battery testing. The circuit is powered by a 4.5-V battery and compares the battery under test with an internally derived reference, set by R1/R2/P1. Each LED of the 10 represent 10% of full scale. For best results, the battery (D.U.T.) should be loaded with an appropriate resistor.

11

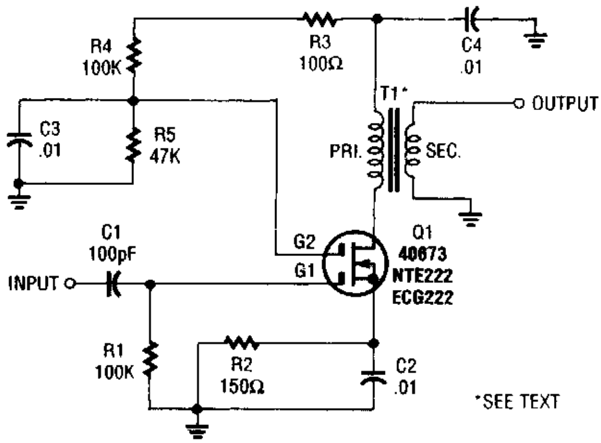
Buffer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

BUFFER/AMPLIFIERS



A

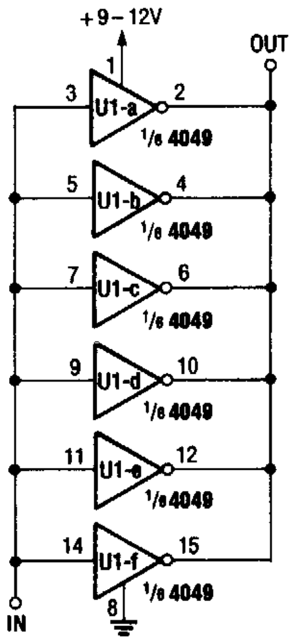


B

POPULAR ELECTRONICS ***FIG. 11-1***

These two buffer/amplifiers that have been successfully used with VFOs: one (shown in A) is based on a pair of bipolar npn transistors, and the other (shown in B) is built around a dual-gate MOSFET.

HIGH CURRENT BUFFER

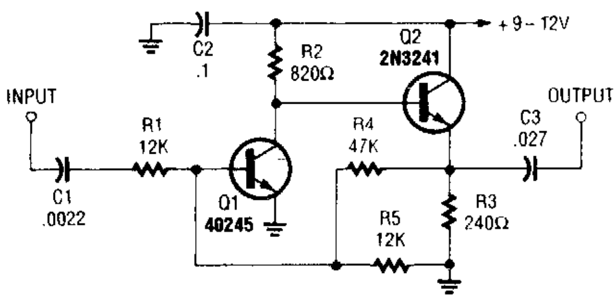


By parallel connecting all six gates of this 4049 hex inverting buffer, you can obtain a much higher output current than would otherwise be available.

POPULAR ELECTRONICS

FIG. 11-2

VFO BUFFER AMPLIFIER

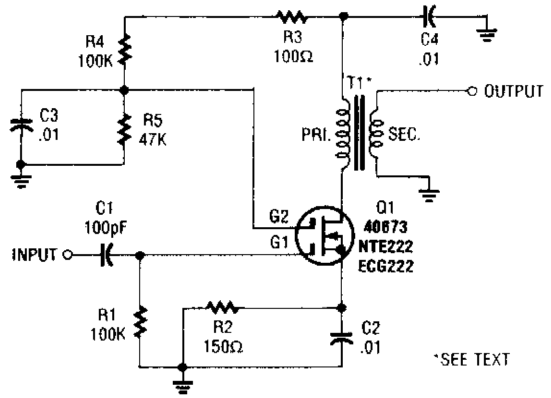


POPULAR ELECTRONICS

FIG. 11-3

A two-transistor feedback pair provides broadband operation. The gain is approximately R_4/R_1.

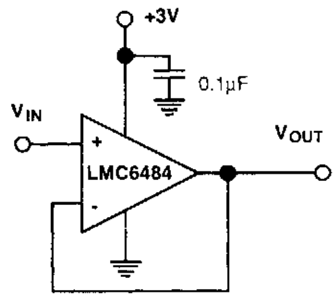
MOSFET BUFFER AMPLIFIER



POPULAR ELECTRONICS **FIG. 11-4**

A MOSFET is used as a wideband buffer amplifier. T1 is wound on a toroid of approximately ½" diameter, with material suitable for frequency (usually 1- to 20-MHz range). The turns ratio should be about 4:1 depending on load impedance. Typically, at 4 MHz, there are 18 turns on the primary, 4 turns on the secondary, and the stage gain is about 14-dB voltage ($Z_L = 50\ \Omega$).

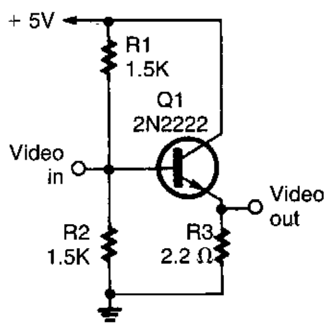
3-V RAIL-TO-RAIL SINGLE-SUPPLY BUFFER



NATIONAL SEMICONDUCTOR **FIG. 11-5**

The LMC6484 provides a 3-V p-p rail-to-rail buffer with a +3-V supply commonly used for logic systems.

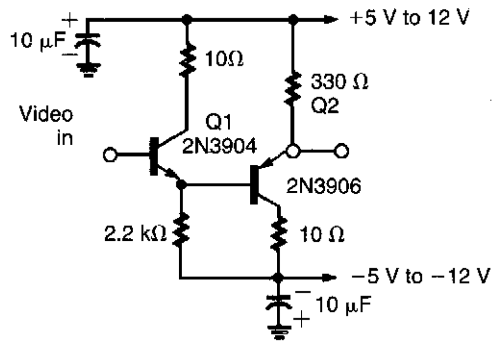
SIMPLE VIDEO BUFFER



ELECTRONICS NOW **FIG. 11-6**

This simple emitter follower can be used as a video buffer.

LOW-OFFSET SIMPLE VIDEO BUFFER



WILLIAM SHEETS **FIG. 11-7**

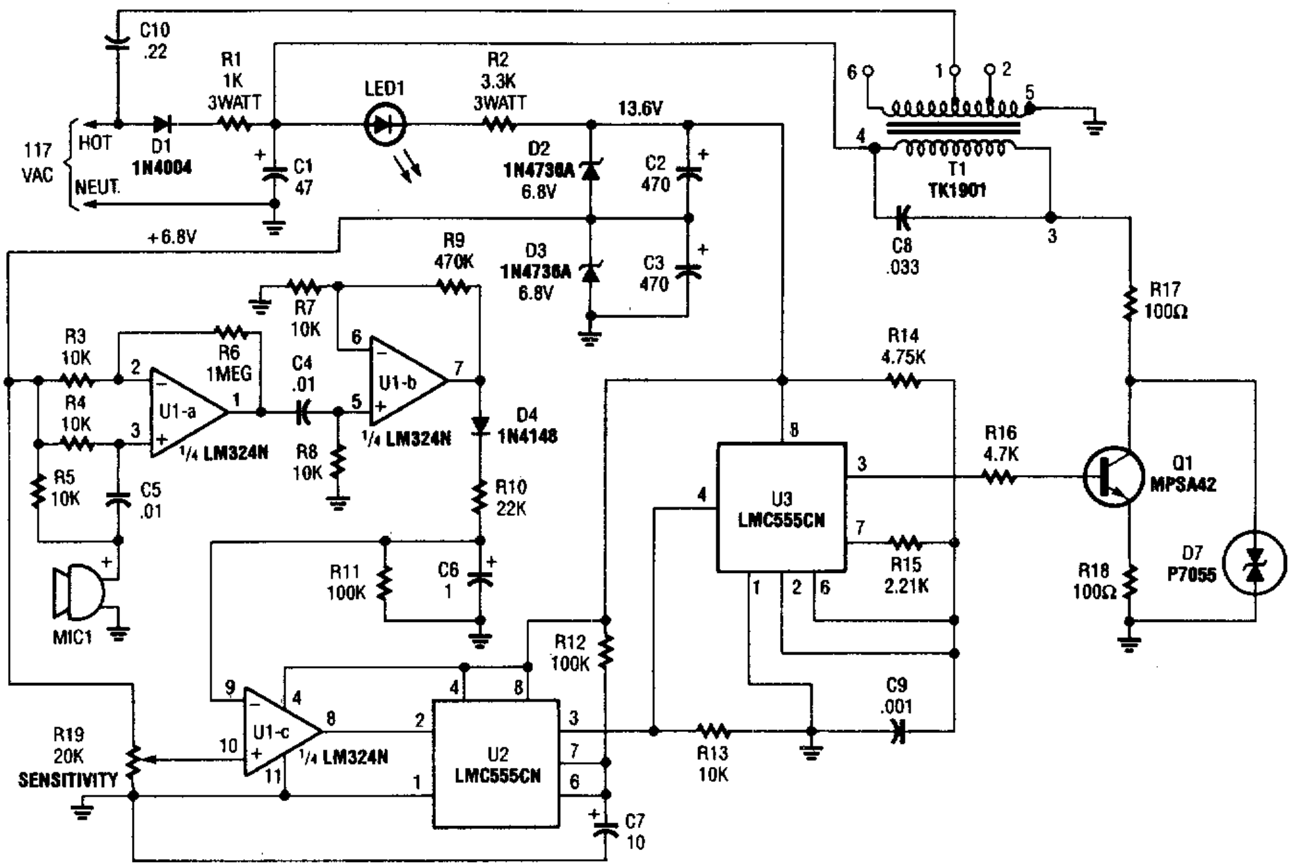
This circuit has proved to be an effective video buffer and will easily drive a 75-Ω load to 1.5-V p-p output. BW is better than 20 MHz and there is less than 0.05-V dc offset, which is the difference in V_{BE} of Q1 and Q2. The supply lines should be well bypassed, ± 5 V or more.

12

Carrier-Current Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

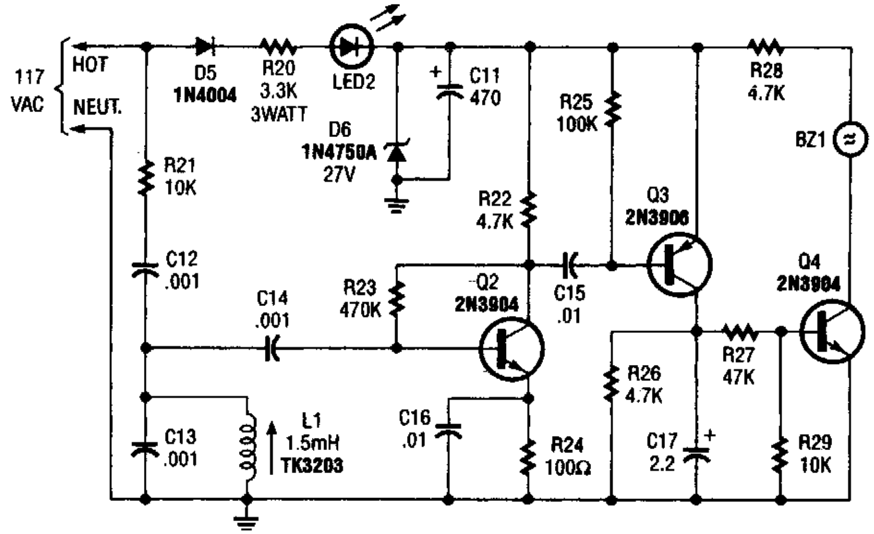
CARRIER-CURRENT BABY-ALERT TRANSMITTER



1993 ELECTRONICS HOBBYIST HANDBOOK *FIG. 12-1*

The baby-alert transmitter is built around an LM324 quad op amp (U1), two LMC555CM CMOS oscillator/timers (U2 and U3), and a few support components. The transmitter sends a signal on receipt of a sound at MIC1. It has a frequency of around 125 kHz and can be used to trigger an alarm receiver.

CARRIER-CURRENT BABY-ALERT RECEIVER



1993 ELECTRONICS HOBBYIST HANDBOOK

FIG. 12-2

The baby-alert receiver is comprised of three transistors: Q2, which is configured as a high-gain linear amplifier; Q3, which serves as both an amplifier and detector; and Q4, which is essentially used as a switch; and a few additional components. It sounds an alarm BZ1 on receipt of a 125-kHz signal from an alarm transmitter via the 120-V power lines.

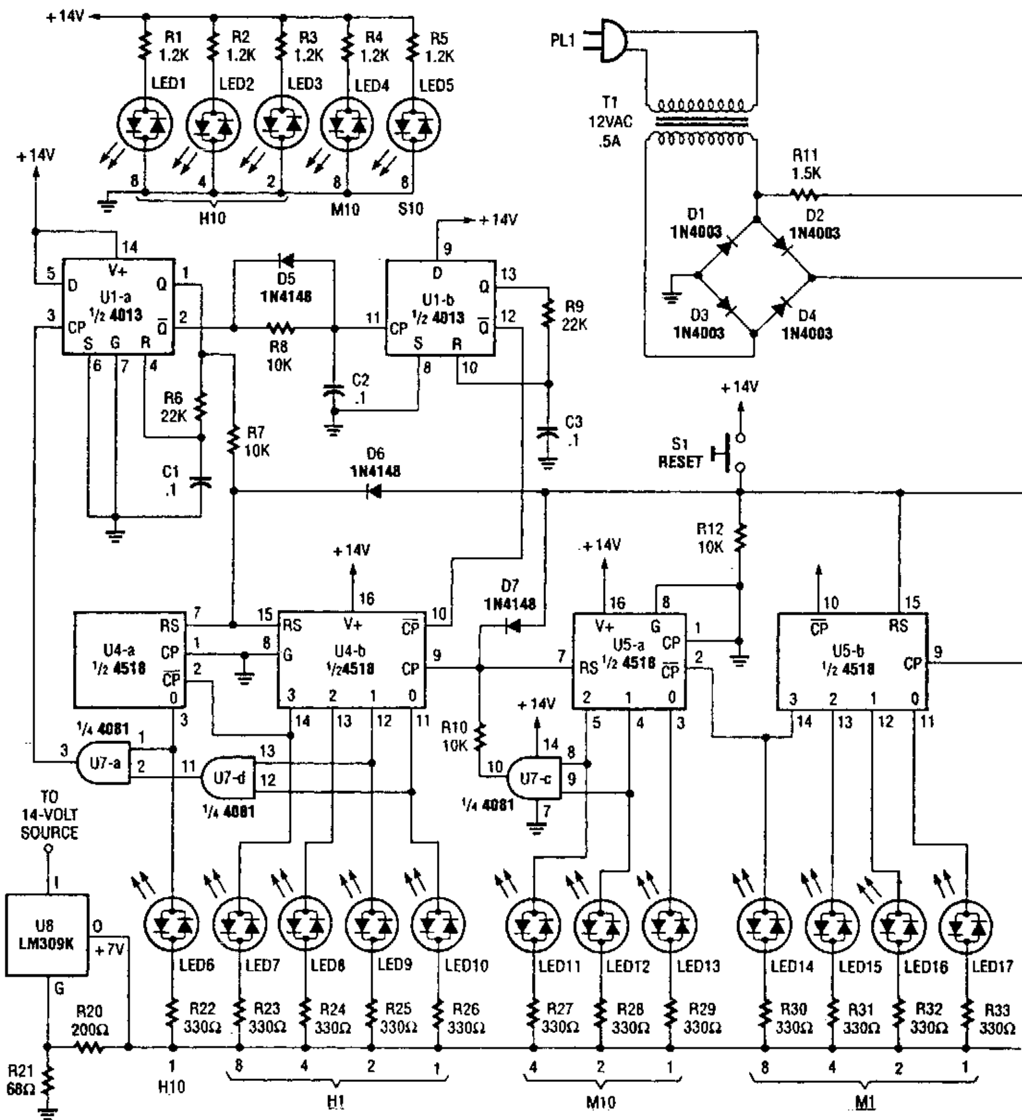
13

Clock Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Binary Clock

BINARY CLOCK



1992 PE HOBBYIST HANDBOOK

This circuit is an unusual clock in that the LEDs are bi-color red/green displays that indicate the time in binary coded decimal form.

LEDs 21 through 24 read out seconds

LEDs 5, 18, 19, and 20 read out 105 seconds

LEDs 14 through 17 read out in minutes

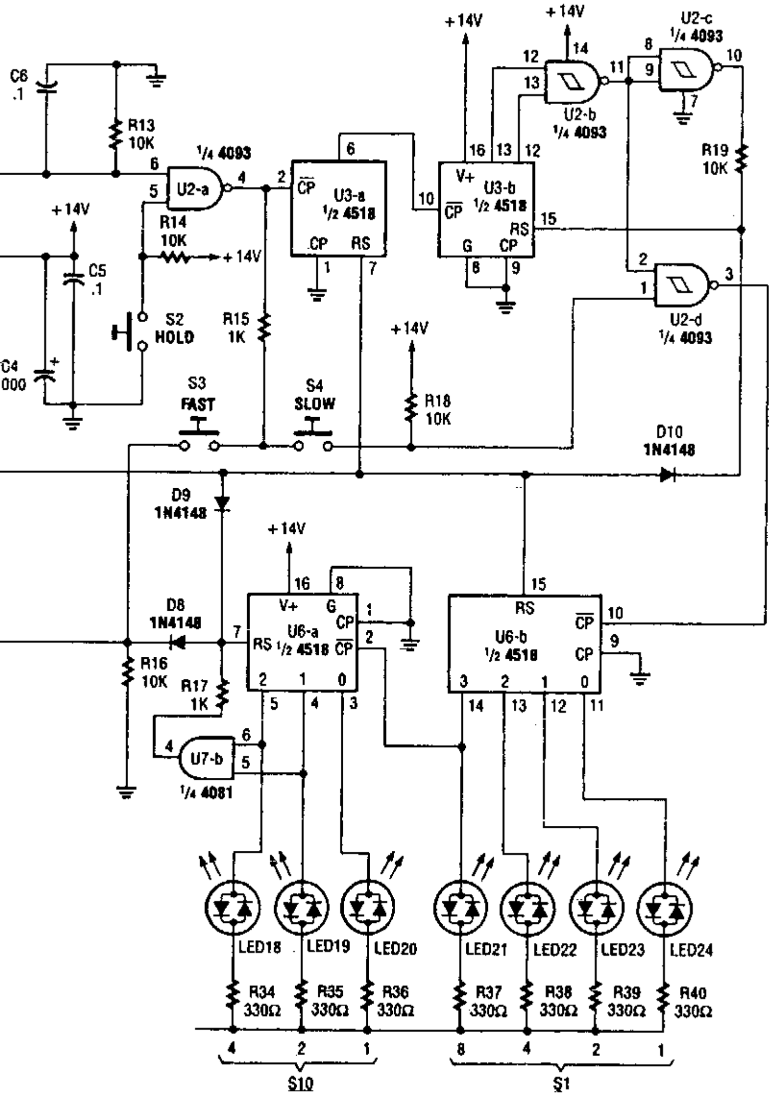


FIG. 13-1

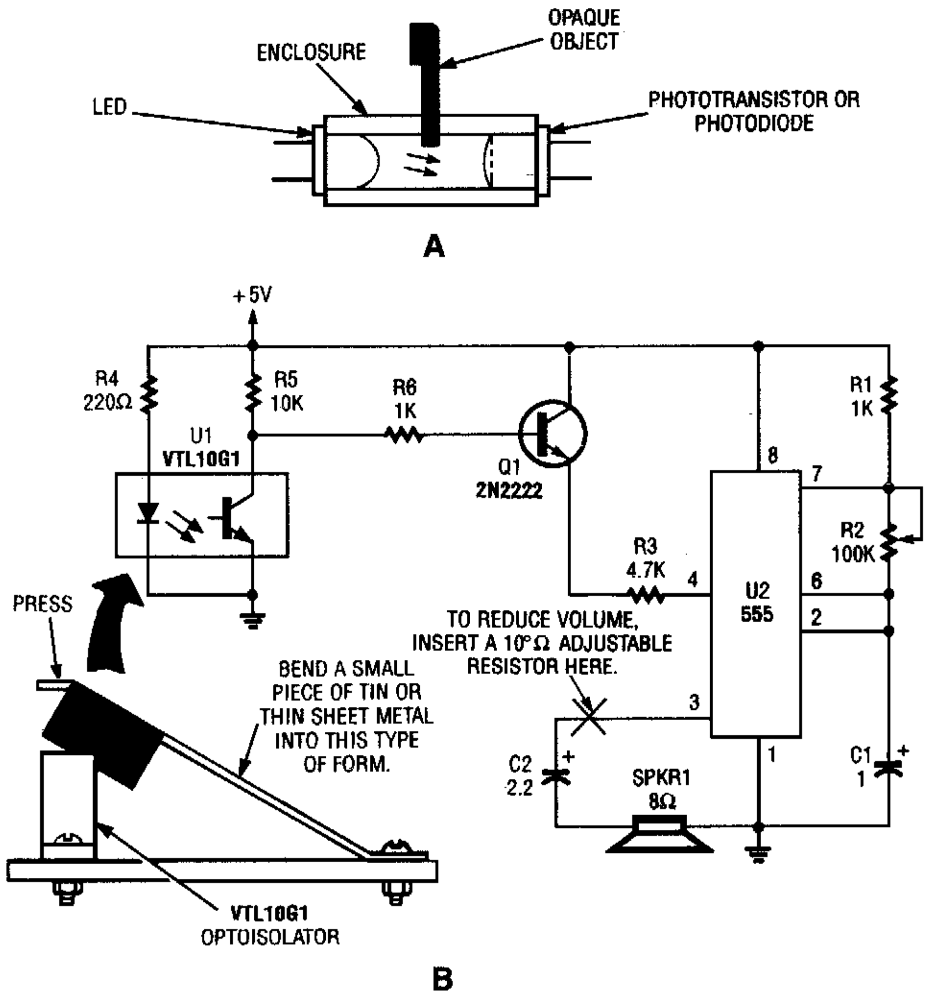
LEDs 4, 11, 12, and 13 read out in 105 minutes
LEDs 7 through 10 read out the hours
LEDs 1, 2, 3, and 6 read out tens of hours
The 60-Hz line is used as a timebase.

14

Code Practice Circuits

The source of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

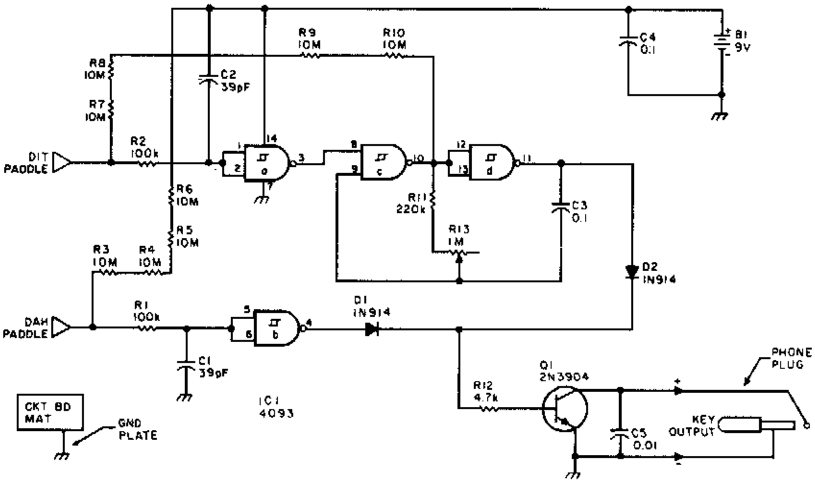
CODE PRACTICE OSCILLATOR USES OPTOISOLATOR



POPULAR ELECTRONICS *FIG. 14-1*

A slotted-pair isolator (A) is effectively an enclosed-pair isolator with a slit that will allow an obstacle to interrupt the light path. That could be useful for building a code key (B).

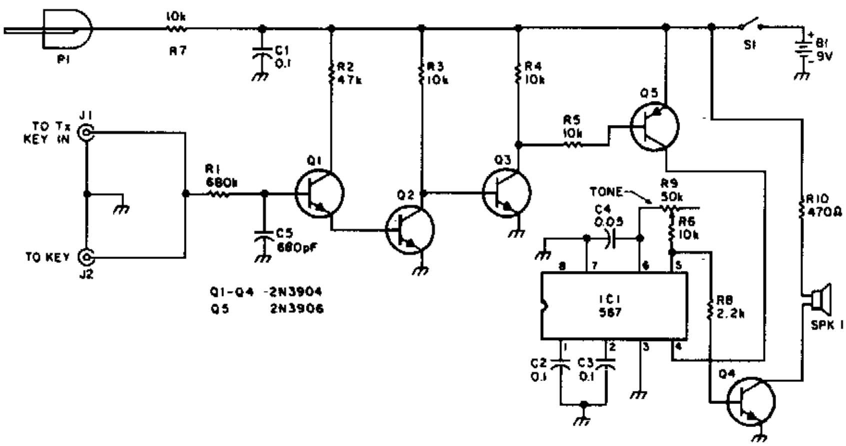
ELECTRONIC CW "BUG" KEYER



73 AMATEUR RADIO TODAY **_FIG. 14-2_**

This keyer uses skin conductivity to simulate the old-fashioned mechanical CW bug keyer. When the "dit" paddle is touched the bias on the inverter, IC1-a is shunted to ground, and it produces a logic high, causing oscillator sections C&D to generate a low-frequency square wave keying Q1 for a series of "dits." When the "dah" paddle is touched, section b produces a logic high, driving keyer Q1 on.

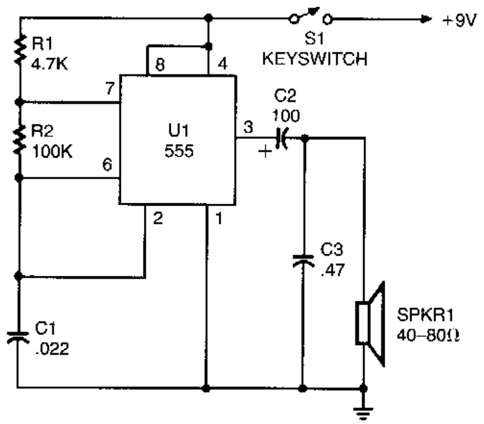
QRP SIDETONE GENERATOR/CODE PRACTICE OSCILLATOR



73 AMATEUR RADIO TODAY **_FIG. 14-3_**

For use with low-power transmitters with a positive keying voltage. Q1/Q2/Q3 form a switching amplifier. When the key is pressed, the collector of Q3 goes to ground, turning on Q5 and activating IC1, an audio oscillator. Q4 drives the speaker. For use as a code practice oscillator, insert P1 and J1 and a key in J2.

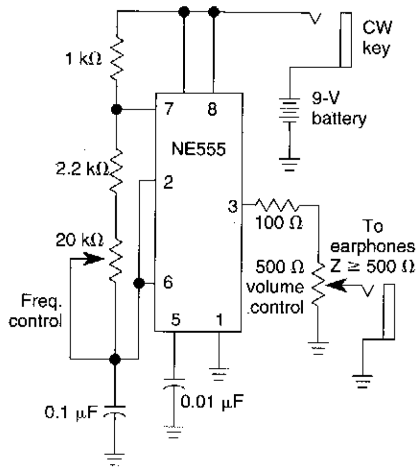
MORSE PRACTICE OSCILLATOR



POPULAR ELECTRONICS ***FIG. 14-4***

A 555 timer configured as an astable multivibrator is used in this circuit to generate an audio note. C1 can be changed to vary the audio note as desired.

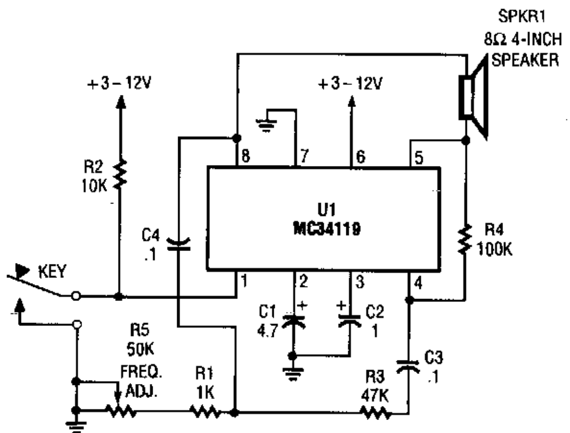
CODE PRACTICE OSCILLATOR



WILLIAM SHEETS ***FIG. 14-5***

The tone and volume of the sound produced when the telegraph key is depressed can be varied in this code practice oscillator.

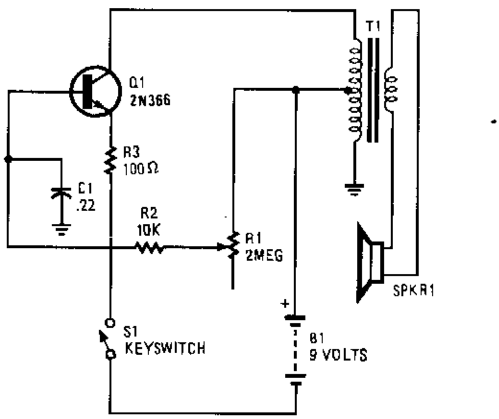
VARIABLE FREQUENCY CODE PRACTICE OSCILLATOR



POPULAR ELECTRONICS ***FIG. 14-6***

The variable frequency audio oscillator can be used as a low-level alarm sounder or a code-practice oscillator.

SINGLE-TRANSISTOR CODE PRACTICE OSCILLATOR



POPULAR ELECTRONICS ***FIG. 14-7***

A 2N366 is configured as an audio feedback oscillator using an audio transformer is shown. Adjust R1 for proper operation and desired audio note.

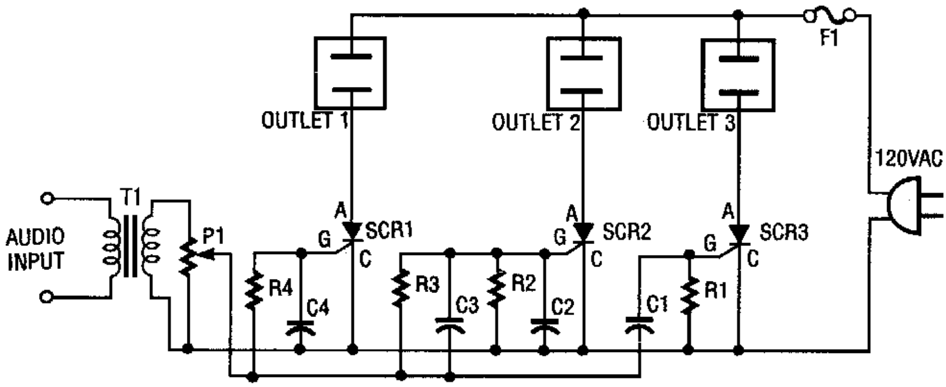
15

Color Organ Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

3-Channel Color Organ

3-CHANNEL COLOR ORGAN



1991 PE HOBBYIST HANDBOOK ***FIG. 15-1***

The ac line power is brought back into the circuit through F1, a protective 5-A fuse. One side of the ac line is connected to one side of each ac outlet. The other side of the ac line is connected to each SCR or silicon-controlled rectifier. Each SCR is, in turn, connected to the other side of each ac outlet.

An audio signal is brought into the circuit from a stereo speaker by transformer T1. This transformer has 500-Ω impedance on the primary and 8-Ω impedance on its secondary. Connect T1 so that the 8-Ω side is connected to the speaker and the 500-Ω side is connected to potentiometer P1.

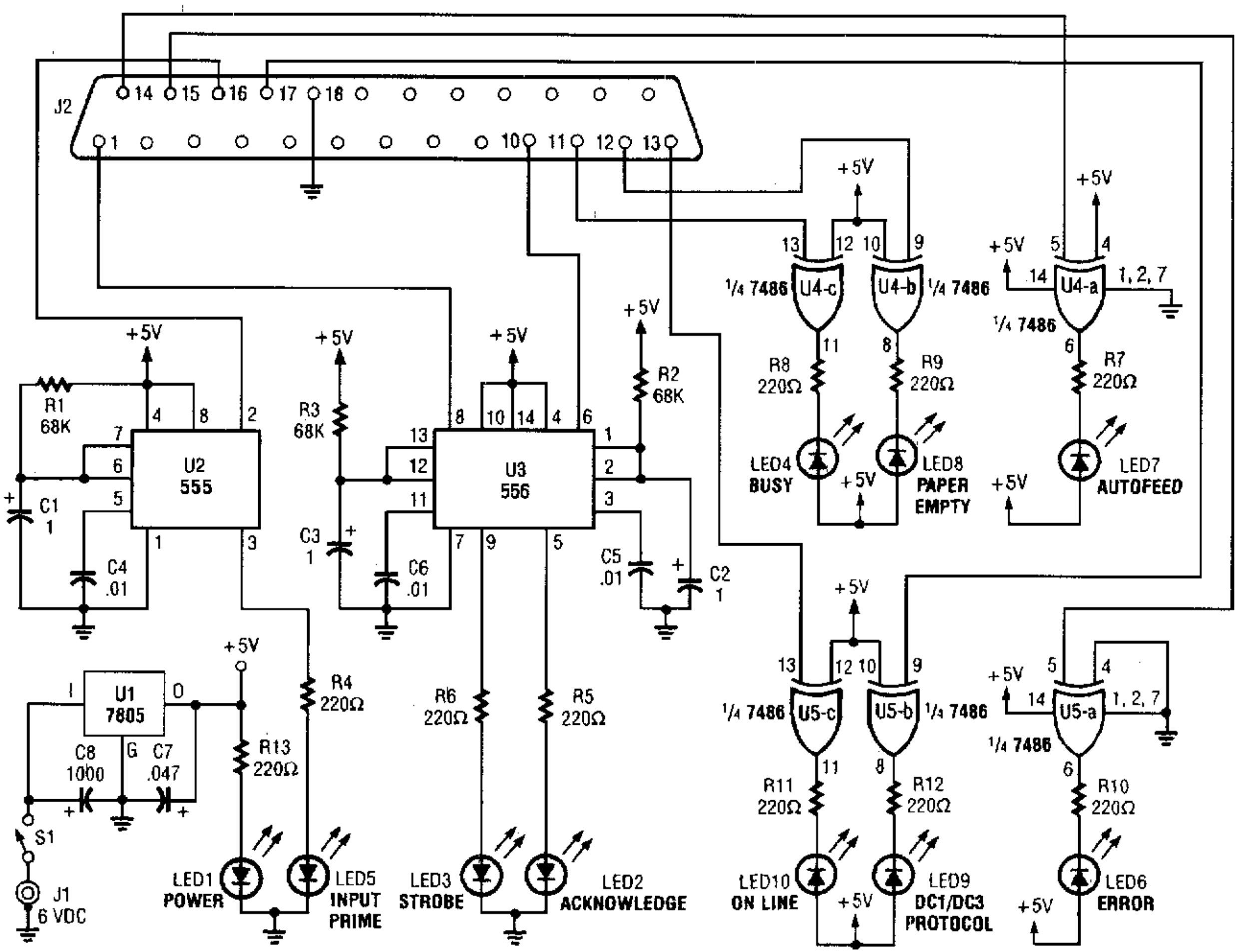
Potentiometer P1 is used as a level or sensitivity control. The signal from its wiper lead is applied to each RC filter stage. Because each SCR has a different RC (resistor/capacitor) filter on its gate lead, each will respond to different frequencies. The greater the capacitance in the filter, the lower the frequency that the SCR will respond to.

16

Computer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

PRINTER SENTRY



POPULAR ELECTRONICS

FIG. 16-1

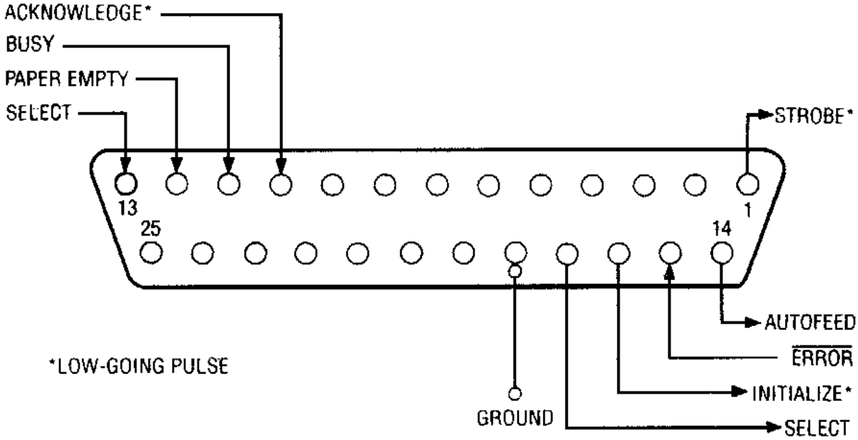
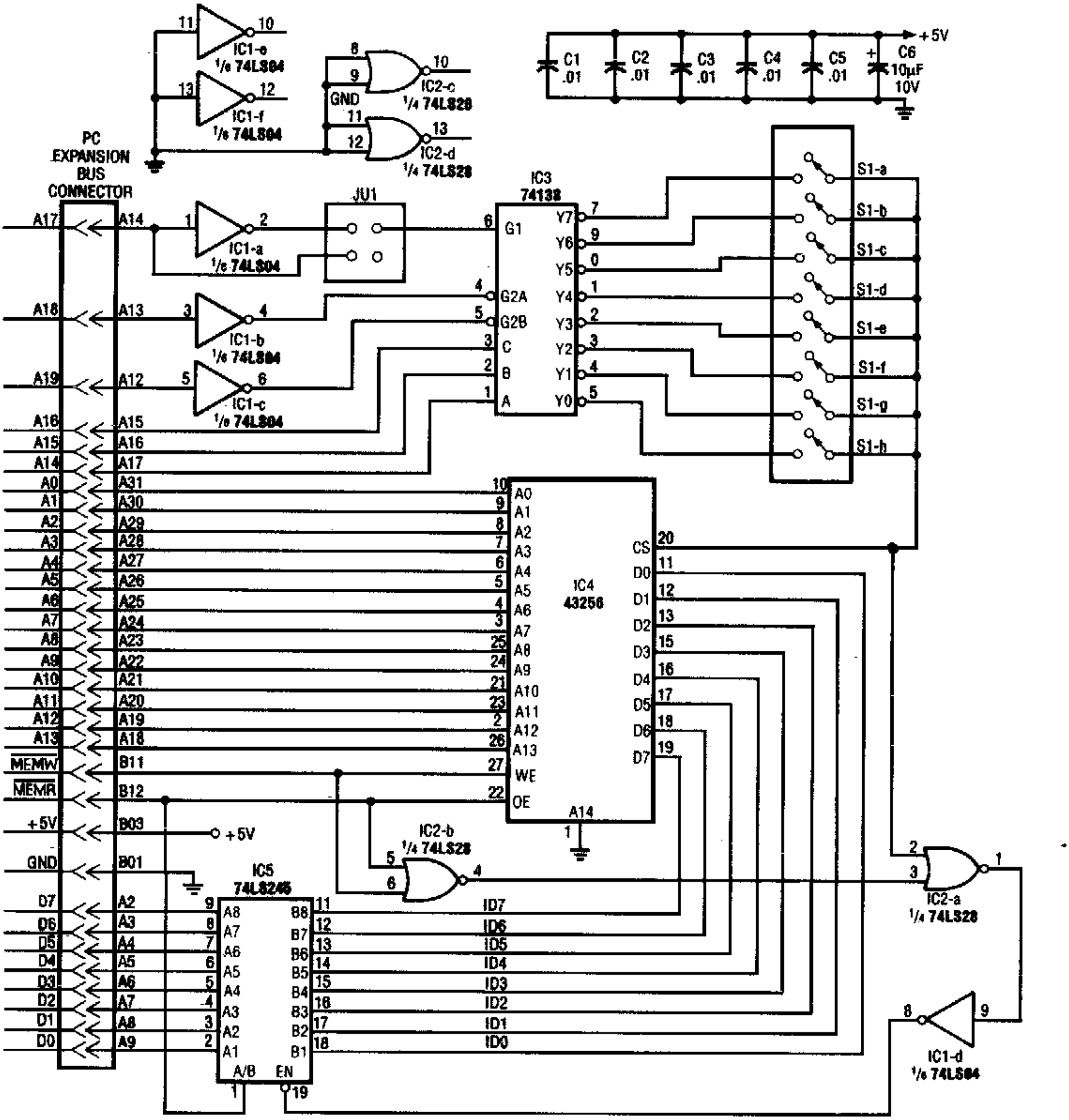


TABLE 1—PIN CORRESPONDENCE

DB-25 Connector	Centronics-Style Connector
1	1
10	10
11	11
12	12
13	13
14	14
15	32
16	31
17	36
18	19

Handy for monitoring printers, this circuit displays all the signals on a parallel link. It monitors the status of the lines, enabling remote monitoring of the operation of a printer, and it also gives an indication of troubles (paper empty, busy, etc.).

PC PASSWORD PROTECTION

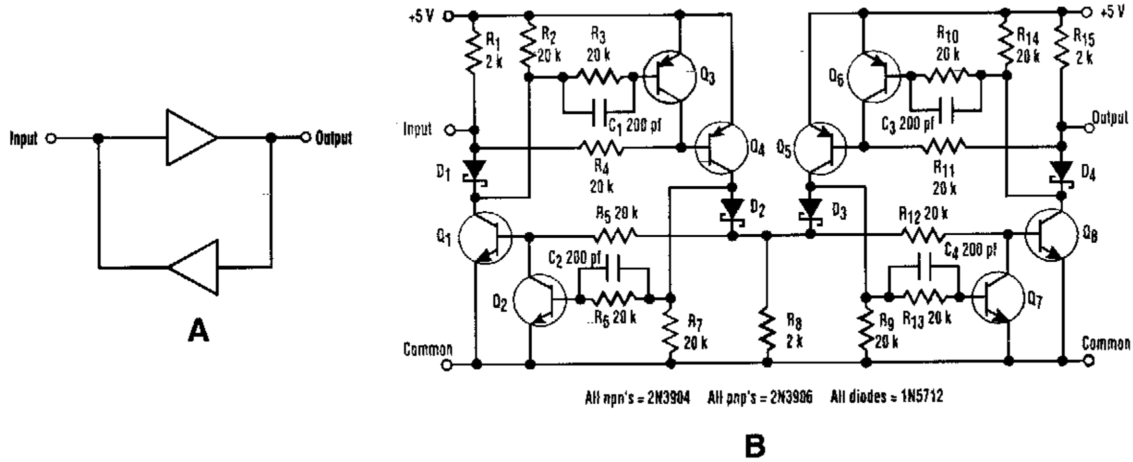


ELECTRONICS NOW

FIG. 16-2

With this circuit, a PC will be protected, requiring a password to boot. After three times, the computer will have to have a cold reboot and the password tried again. Software for this system is available—consult the reference for further details.

BUFFER I²C DATA AND CLOCK LINES



ELECTRONIC DESIGN

FIG. 16-3

The I²C serial bus is a popular two-wire bus for small-area networks. I²C Clock and Data lines have open collector (or drain) outputs for each device on the network. Only a single pull-up resistor is needed. With this architecture, each device can "talk" on the network, rather than just "listen." In some circumstances, it might be desirable to buffer these lines to expand the network, which can sometimes be a tricky task. The obvious approach (Fig. 1) wont work because it latches in either the higher or lower state. A circuit for a noninventory nonlatching buffer is also shown.

The circuit is symmetrical about its center so that the input and output can be swapped. Q1 and Q8 are the output open collector drivers. Q2, Q3, Q6, and Q7 provide the nonlatching functions. The capacitors prevent switching glitches by ensuring the inhibit transistors turn off before the output transistors do.

Operation can be best explained by example: if the input is high, Q4 turns off, and the voltage across R8 goes to zero. This turns off Q1 and Q8. The output then goes high, which is the circuit's normal resting place. If the input is pulled low, Q4 is turned on.

Diode D1 remains reverse-biased, preventing Q3 from turning off Q4. With Q4 on, current is supplied to both Q2 and Q1 to turn them on, but Q2 turns on first to keep Q1 off. This prevents the input from latching. Q4 also turns on Q8. D4 is now forward-biased, so Q6 turns on, and thus turns off Q5. With Q5 off, Q7 will not turn on. The output remains low. Even with both the input and the output externally driven low, the circuit will not latch. The circuit, using the values shown in Fig. 2, reached a clock rate of 80 kHz with a VOH of 5.0 V and a VOL of 0.5 V.

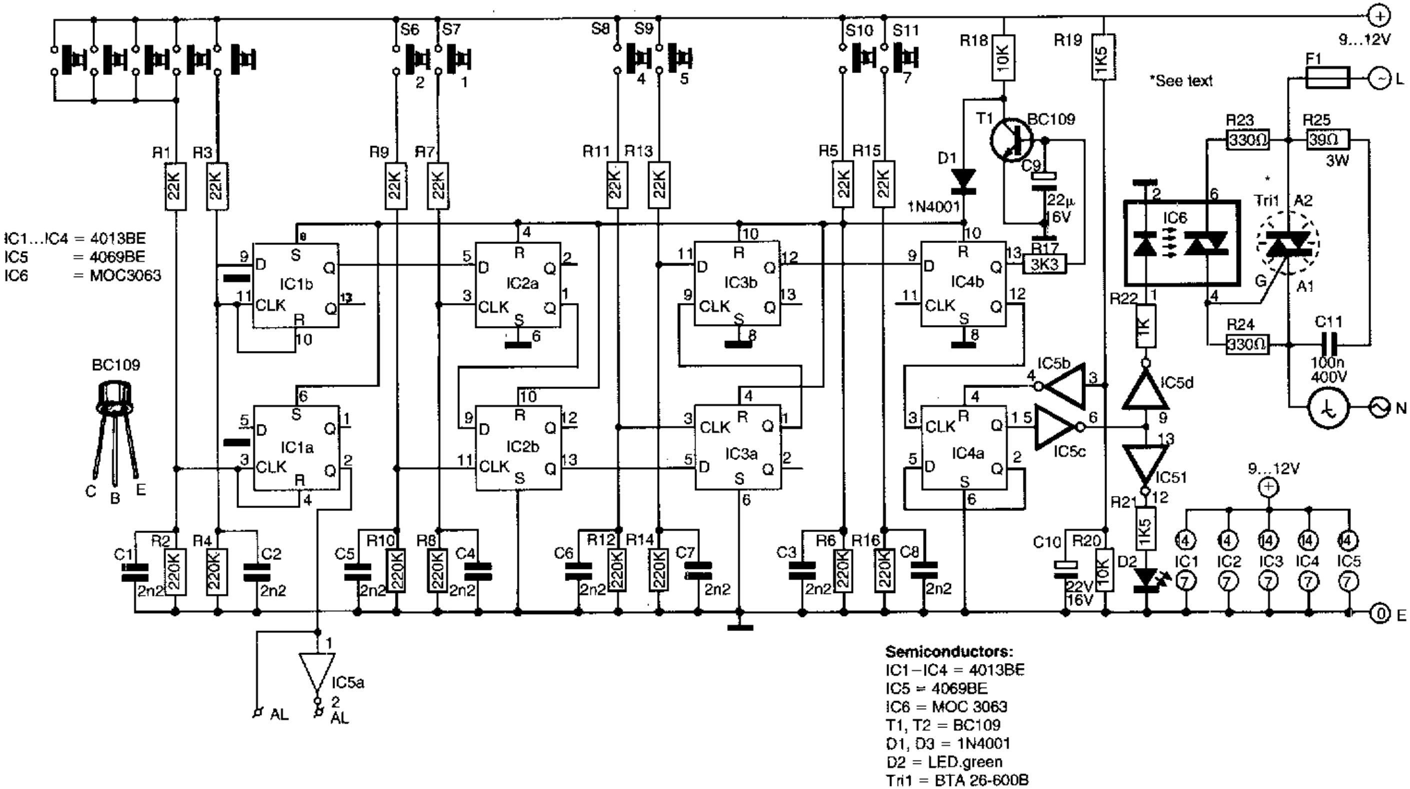
17

Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

6-Digit Coded ac Power Switch
VCR TV On/Off Control
Simple Power Down Circuit
Simple ac Voltage Control
Dual-Control Switch Uses ac Signals

6-DIGIT CODED ac POWER SWITCH

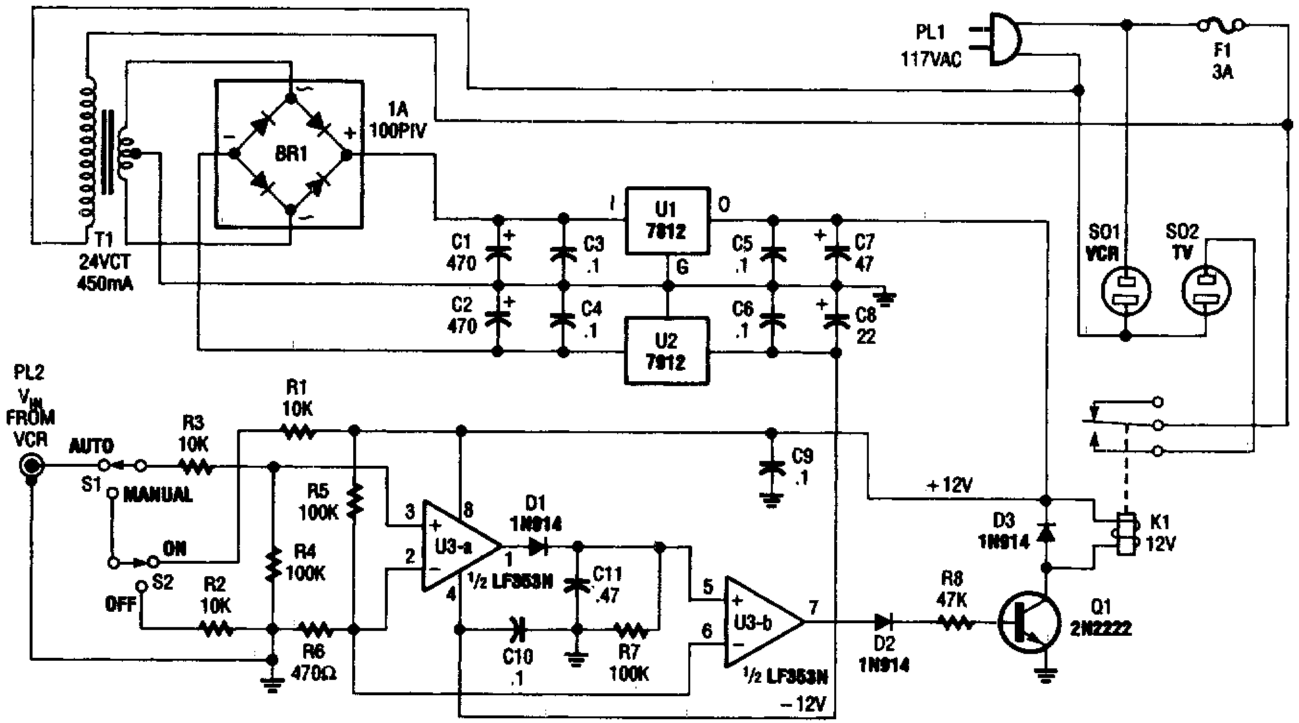


ELEKTOR ELECTRONICS

FIG. 17-1

This switch uses four CD4013 BE dual flip-flops, an inverter, and an optoisolator to drive a triac. The circuit can switch 25-A ac load current. A standard 4 × 3 telephone keyboard is used to enter a 6-digit code. In case of a wrong code, a signal is available to activate an alarm. The disarming method is a secret reset button that can be any number on the keyboard.

VCR TV ON/OFF CONTROL

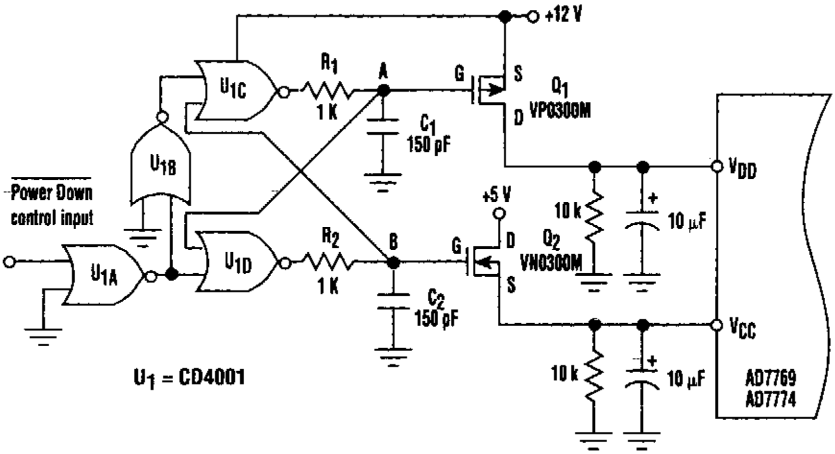


1993 ELECTRONICS HOBBYISTS HANDBOOK

FIG. 17-2

This circuit senses the video from the VCR. When the VCR is turned on, video signal is amplified by U3A and B to drive Q1, activating K1. In this manner, it is not necessary to turn on and off two video devices every time. In many cases, this avoids the use of a cable box, the cable-ready VCR performing this function.

SIMPLE POWER DOWN CIRCUIT



ELECTRONIC DESIGN

FIG. 17-3

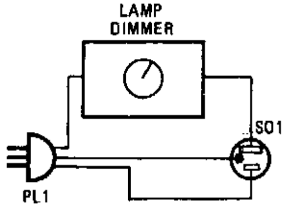
This circuit adds a power-down function to analog I/O ports (for example, the AD7769 and AD7774). Moreover, the diodes ordinarily needed to protect the devices against power-supply missequencing can be eliminated (see the figure).

In the circuit, MOSFETs Q1 and Q2 switch the +5- and +12-V supplies, respectively, in a sequence controlled by two cross-coupled CD4001 CMOS NOR gates (U1C and U1D). The sequence in which power is applied is important: The controlled circuits may be damaged anytime V_{CC} exceeds V_{DD} + 0.3 V. Consequently, the NOR gates must be powered from a 12-V supply throughout the power-down sequence.

Bringing the power down control high (+5 V) applies power to the controlled circuit by turning on all MOSFETs. Specifically, raising the power-down brings the output of U1C low, causing capacitor C1 to discharge VOL exponentially with time constant R_1C_1. As the voltage on C1 falls, two events occur. First, it puts a negative gate-source voltage on P-channel Q1, turning it on.

Second, it causes output gate U1D to go high. With the output of U1D high, capacitor C2 charges exponentially to VOH—about 12-V—applying a positive gate-source voltage to turn on Q2. In the power down mode, the Power Down control is brought low and the RC circuits and their delays work in reverse. Consequently, capacitor C2 discharges to the logic input of U1C before C1 can charge. Hence, Q2 turns off before Q1.

SIMPLE ac VOLTAGE CONTROL

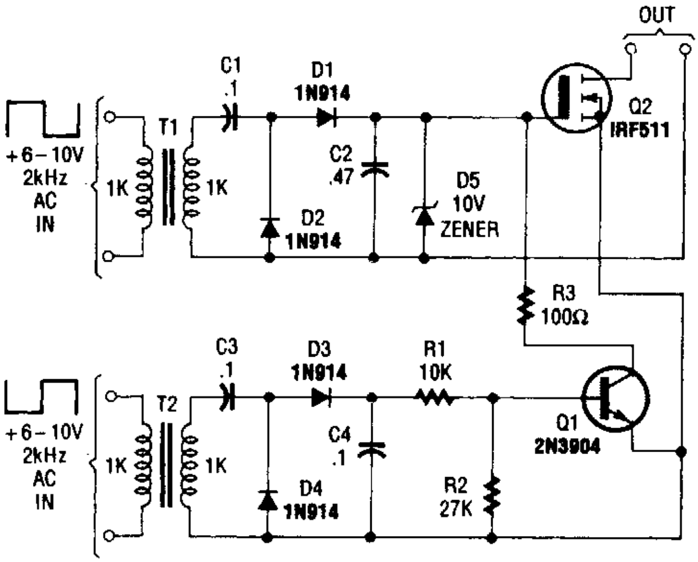


Lamp dimmers can be used for more than just controlling lights. Just provide one with an ac line cord and a socket, and discover just how useful they can be.

POPULAR ELECTRONICS

FIG. 17-4

DUAL-CONTROL SWITCH USES ac SIGNALS



POPULAR ELECTRONICS

FIG. 17-5

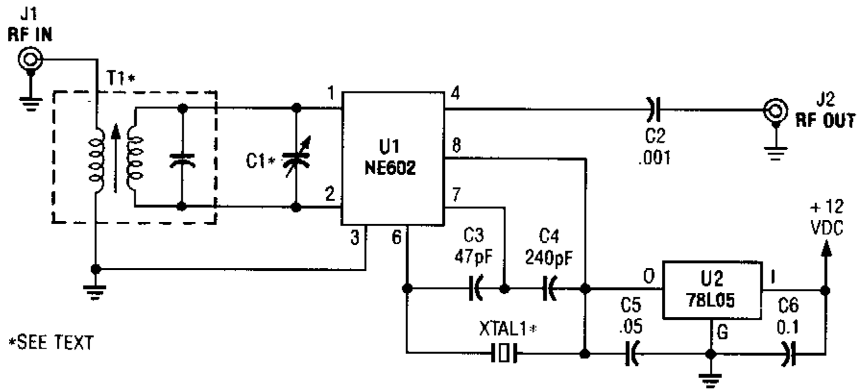
The Dual-Control Switch uses two 6–10-Vac sources to trigger the circuit on and off; one source for each function.

18

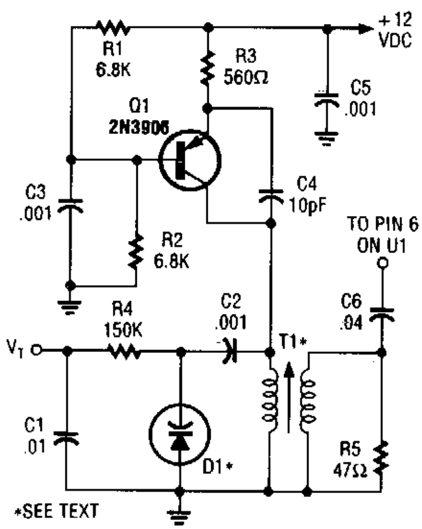
Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ONE-CHIP CRYSTAL-CONTROLLED CONVERTER



A



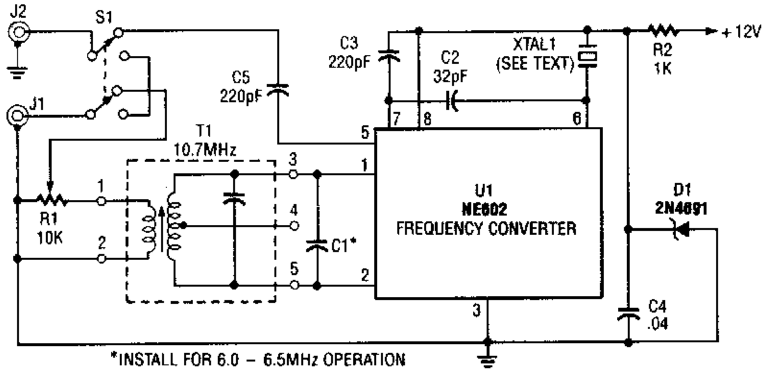
B

POPULAR ELECTRONICS

FIG. 18-1

This circuit can work over a wide range of frequencies. XTAL 1 is a fundamental-frequency crystal. T1 and C1 are tuned to the input frequency. An application of this circuit is a simple shortwave converter for AM radios, etc. A tuneable oscillator can also be used, as shown.

HIGH-PERFORMANCE SHORTWAVE CONVERTER



1990 PE HOBBYIST HANDBOOK ***FIG. 18-2***

The NE602 chip, U1, contains oscillator and mixer stages. The mixer combines the oscillator signal with the input RF signal to produce signals whose frequencies are the sum and difference of the input frequencies. For example, an 8.5-MHz oscillator and a 10-MHz incoming signal will give output signals at 18.5 MHz (10 + 8.5) and 1.5 MHz (10 – 8.5). Recall that 1.5 MHz is 1500 kHz and an ordinary AM radio will tune to it.

The choice of crystal depends on what shortwave band you want to hear. The 9.5- to 10-MHz band is less crowded and includes the time-signal station WWV. For that band, you'll need a crystal of 8.5 to 8.9 MHz. There is no standard microprocessor crystal in that range, but you can use an amateur radio crystal, have a crystal custom-made, or use a CB crystal.

Transformer T1 rejects signals that are outside the band you are interested in. Transformer T1 should pass signals from 9 to 11 MHz and attenuate all others.

The transformer, T1, used in the circuit is a 10.7-MHz IF transformer salvaged from an FM radio. They are fairly easy to obtain new from parts stores and mail-order houses. Most 10.7-MHz IF transformers will tune across the 9.5- to 10-MHz band without modification; all you need to do is turn its tuning slug. To receive the 6.0- to 6.5-MHz shortwave band, you'll have to add a 150-pF capacitor.*

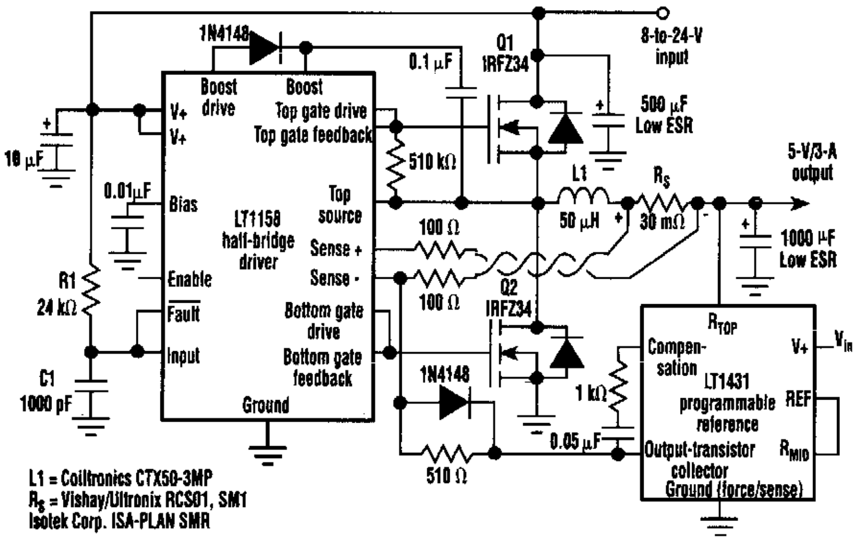
Capacitors

C1	150-pF, ceramic disc (see text)
C2	32-pF, ceramic disc
C3, C5	220-pF, ceramic disc
C4	0.04 or 0.05-µF, ceramic disc

Additional Parts and Materials

U1	NE602N frequency-converter integrated circuit
D1	6.2-V, 0.4 or 1-W Zener diode
R1	10,000-Ω panel-mount potentiometer
R2	1000-Ω, ¼-W, 5% resistor
J1, J2	RCA phono jack
S1	DPDT, toggle switch, panel mount
T1	10.7-MHz IF transformer (green color coded)
XTAL 1	8.5-MHz crystal or CB channel-5 receiving crystal (see text)
XTAL 2	5.0-MHz microprocessor crystal for 6-MHz band

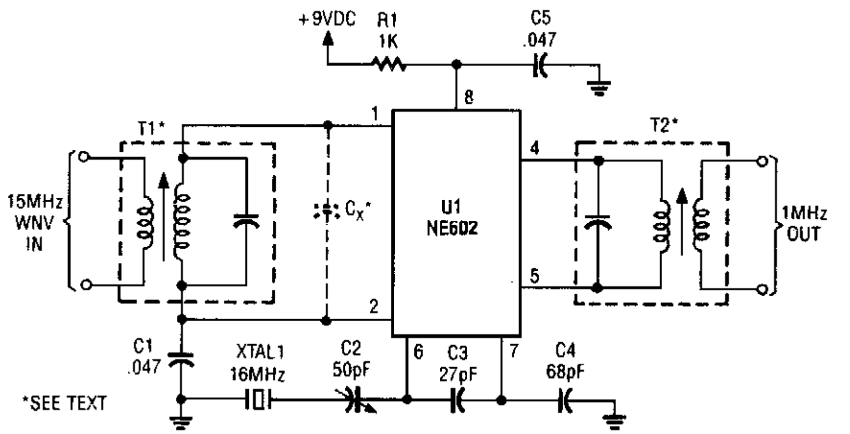
3-A dc-dc CONVERTER NEEDS NO HEATSINK



ELECTRONICS DESIGN *FIG. 18-3*

This regulator delivers 90% efficiency at 12-V input, 5-V output. It uses an LT1158 and LT1431 by Linear Technology, Inc. High efficiency is obtained by synchronously switching two power MOSFETs in a step-down switching regulator. The LT1431 voltage reference combines with the LT1158 half-bridge driver to form a constant off-time current mode loop.

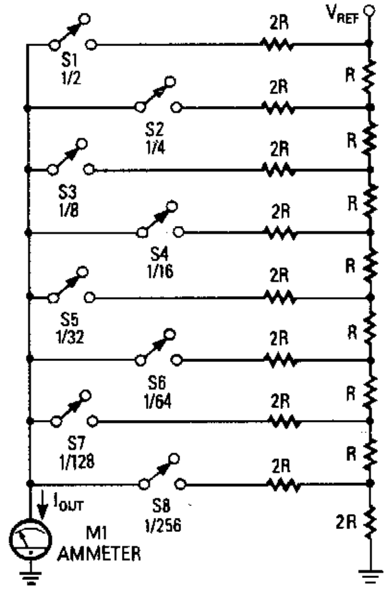
SIMPLE WWV CONVERTER FOR AUTO RADIOS



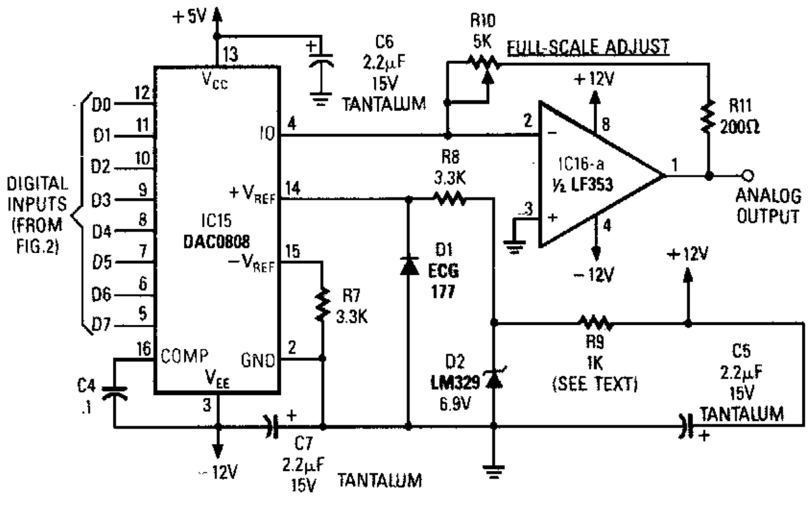
POPULAR ELECTRONICS *FIG. 18-4*

This simple frequency converter mixes the 15-MHz WWV/WVH signal with a 16-MHz signal from the LO to convert it down to 1 MHz so that it can be heard on AM-band receiver.

DIGITAL-TO-ANALOG CONVERTER



A

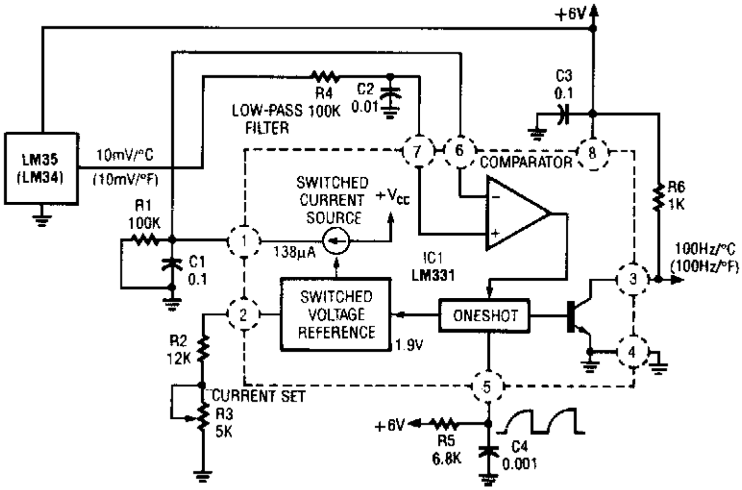


B

1992 R-E EXPERIMENTERS HANDBOOK ***FIG. 18-5***

Figure A is an *R*/2*R* resistor ladder. Each switch that is closed increases the amount of current at I_{out}. A simple channel A/D converter is shown in Fig. B. The voltage reference (D2) is common to all channels, but the value of the dropping resistor (R9) varies as the number of DACs installed in the system. IC15 is a DAC0808 A/D converter chip. ICI6A is an op amp to interface the output current from the D/A convert to an analog voltage output.

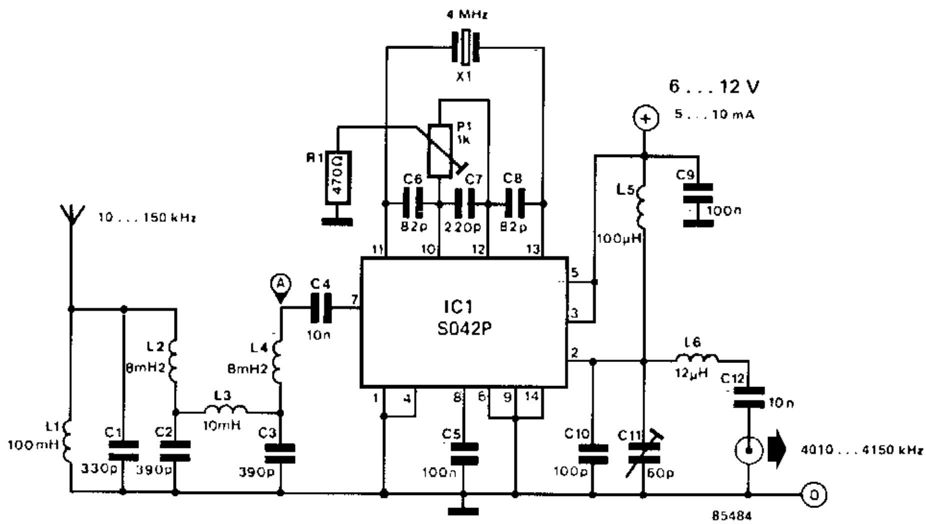
TEMPERATURE-TO-FREQUENCY CONVERTER



RADIO-ELECTRONICS ***FIG. 18-6***

In this circuit an LM34 or LM35 produces a frequency proportional to temperature. Reference current (138 μA) is set via R3. The output can be used to drive a display, frequency counter, or other indicating device for temperature readout.

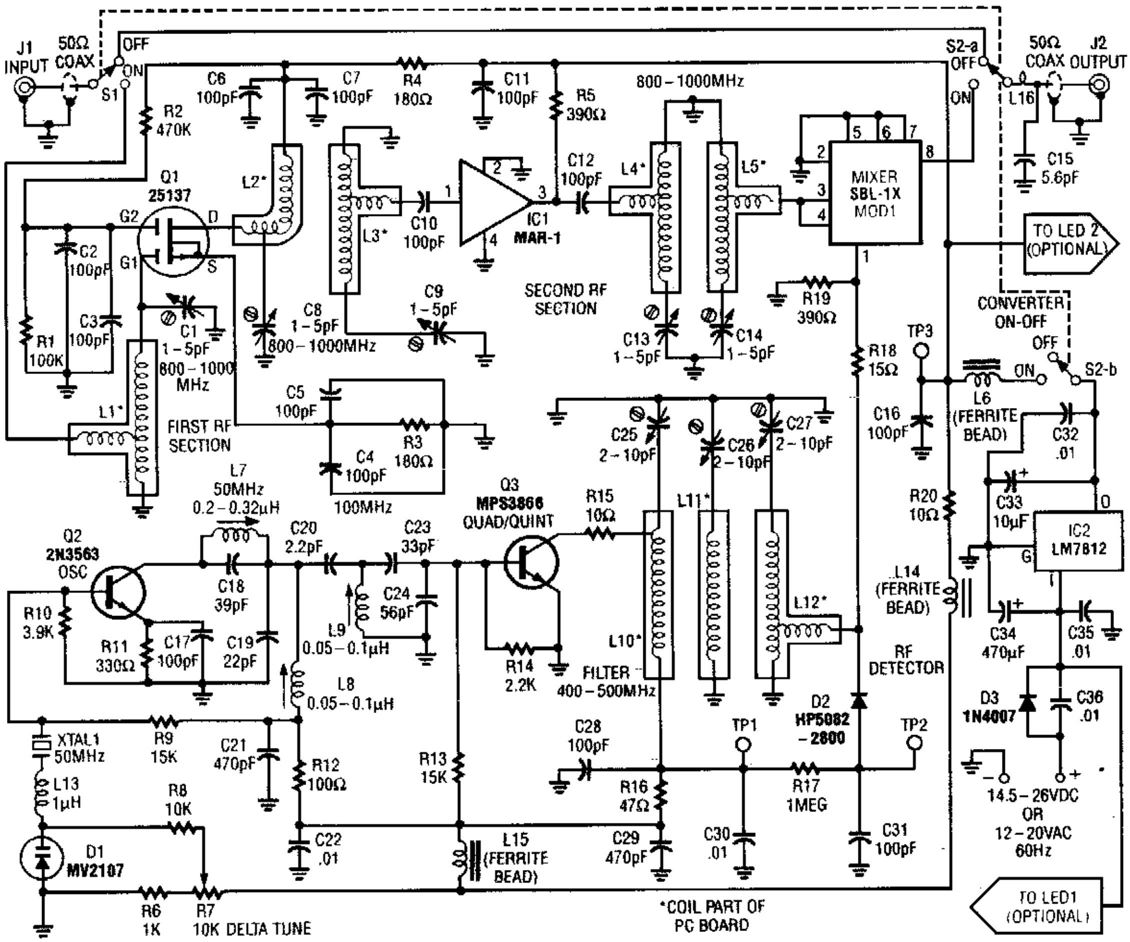
VLF CONVERTER



303 CIRCUITS ***FIG. 18-7***

This converter converts10 kHz to 150 kHz to 4.01 to 4.15 MHz for use with a shortwave receiver for VLF reception. A 4-MHz L.O. frequency is used. X1 can be a microprocessor XTAL or another suitable type. The antenna should be as long as possible.

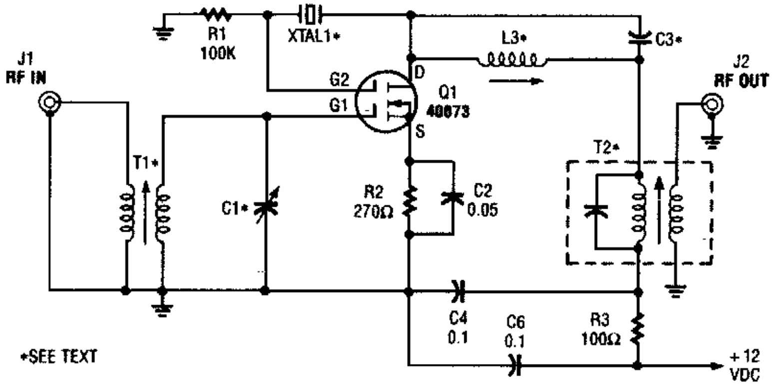
800- TO 1000-MHz SCANNER CONVERTER



RADIO-ELECTRONICS *FIG. 18-8*

This converter enables reception of 800 to 1000 MHz on any scanner covering the 400 to 500-MHz range. The converter can be set up to cover either 800 to 900 MHz or by readjustment 900 to 1000 MHz. Sensitivity is very high because of the GASFET front end. For best results, the scanner should be of a programmable variety. A complete kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

CRYSTAL-CONTROLLED FREQUENCY CONVERTER USING MOSFET

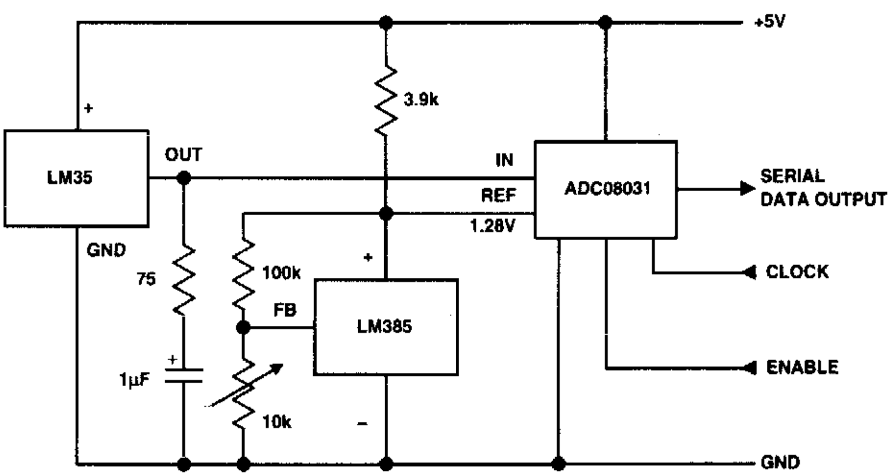


POPULAR ELECTRONICS

FIG. 18-9

The second gate (G2) of a MOSFET can be used to incorporate a crystal oscillator into the same stage as a frequency mixer. Although old hat with tubes, this scheme is seldom seen in dual-gate MOSFET circuitry. L3, C3, and X1 form the crystal oscillator, and T2 is an IF output transformer. T1 and C1 are tuned to the converter input frequency. This circuit should be useable up to 25 MHz or so, or higher with third-overtone crystals.

TEMPERATURE-TO-DIGITAL CONVERTER

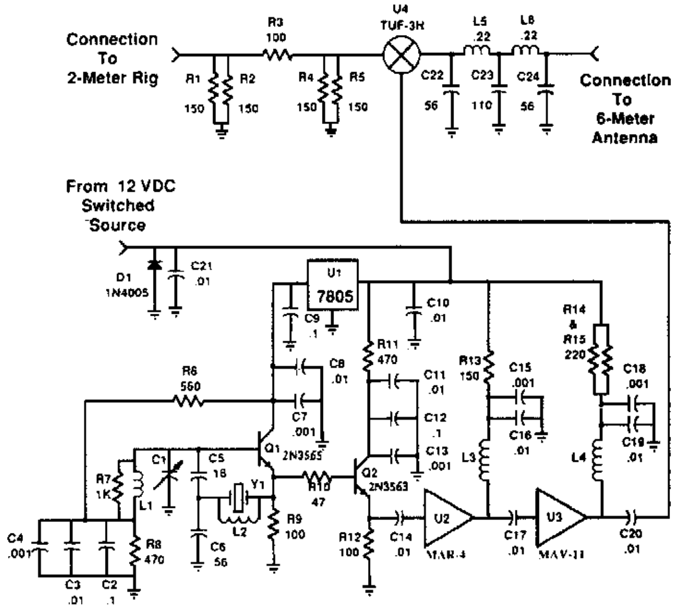


NATIONAL SEMICONDUCTOR

FIG. 18-10

The devices shown from National Semiconductor are used in digital temperature circuit sensor LM35 and reference LM385 feed A-D converter ADC08031.

SIMPLE 2-m–6-m TRANSVERTER

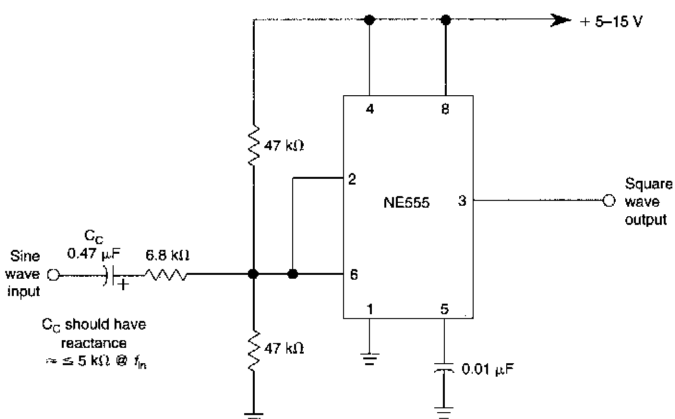


73 AMATEUR RADIO TODAY

FIG. 18-11

Using the bilateral properties of a balanced mixer this transverter will produce 6-m output with 2-m inputs. Y1 is a 90-MHz crystal. Note that the input on 2 m is 143 to 144 MHz for 53 to 54-MHz output. This avoids possibility of extraneous 2-m reception during receive periods. If your radio will not transmit below 144 MHz, then use a 93- or 94-MHz crystal frequency.

SINE- TO SQUARE-WAVE CONVERTER

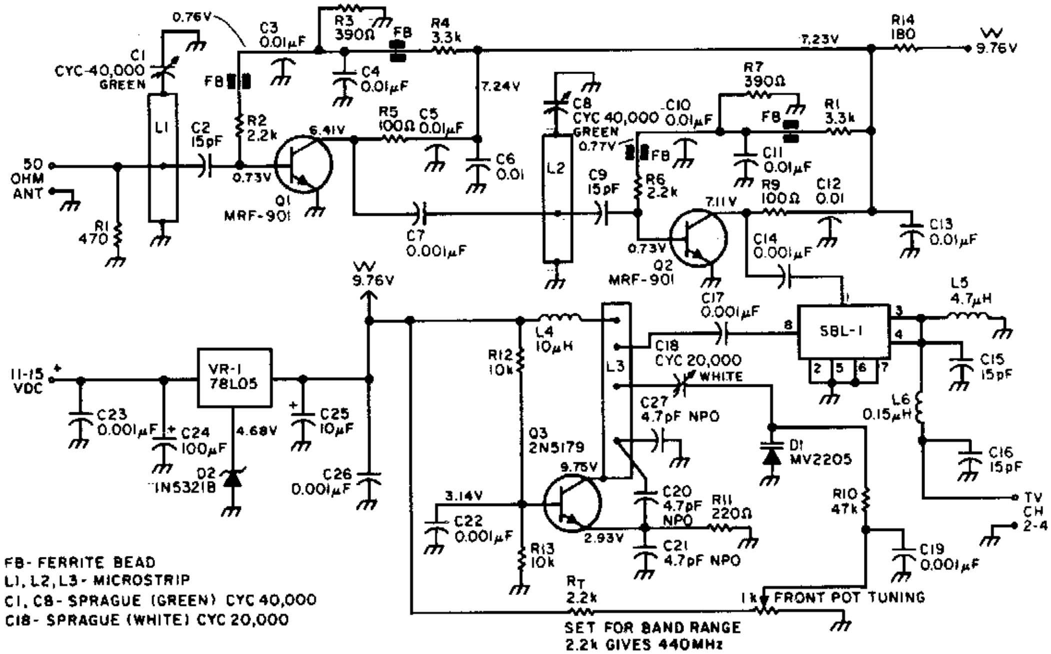


This 555-based Schmitt trigger circuit is useful for creating clock pulses from analog signals since it readily converts sine waves into square waves.

WILLIAM SHEETS

FIG. 18-12

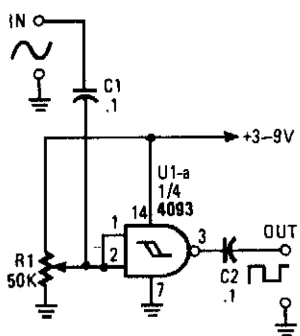
439.25-MHz ATV DOWNCONVERTER



73 AMATEUR RADIO TODAY *FIG. 18-13*

Most ATV (Amateur Television) transmitters transmit a DSB signal and commercial television stations use a VSB (Vestigial Sideband) signal. This fact is made use of in this converter to use the lower sideband. This results in less interference from repeaters that occupy the 440- to 445-MHz portion of the band. However, this approach might suffer from VHF image responses from channel 29, if that channel is active in your area.

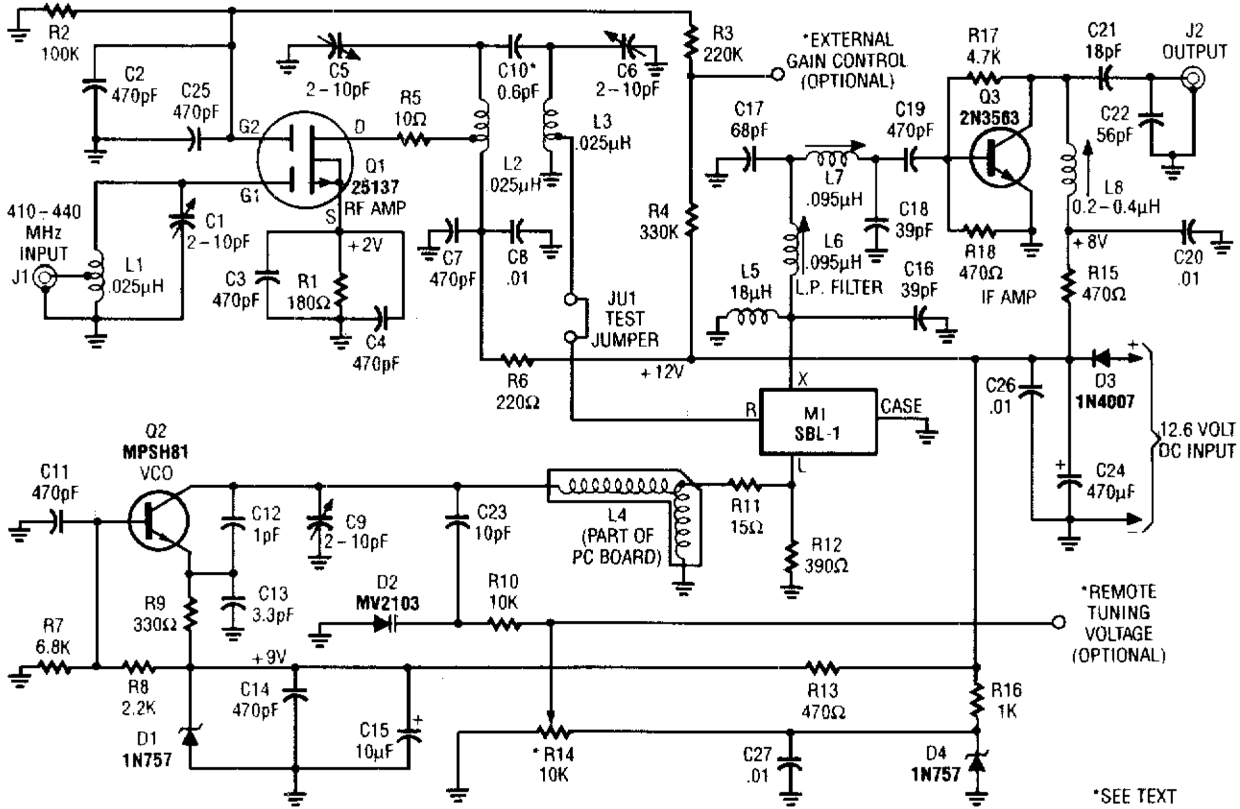
SINE-WAVE-TO-SQUARE-WAVE CONVERTER



This circuit turns a sine wave into a square wave. It is comprised of a single 2-input NAND Schmitt trigger that's configured as an inverter with a trigger level adjustment at its input. As the input voltage rises above the gate's trigger point, the output snaps to its alternate state, producing a square-wave output.

POPULAR ELECTRONICS *FIG. 18-14*

ATV DOWNCONVERTER

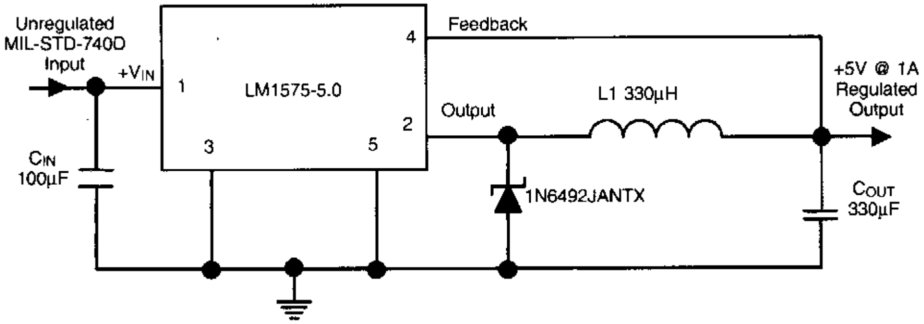


ELECTRONICS NOW

FIG. 18-15

This RF converter converts amateur TV signals in the 420- to 450-MHz region to VHF channel 3 or 4, allowing reception of those signals on a standard TV receiver. RF amplifier Q1 feeds mixer M1, and Q3 acts as an IF amplifier. Q2 is an oscillator operating around 378 MHz and is tuneable over about a 30-MHz range. A complete kit is available from North Country Radio, P.O. Box 53, Wykagyl Station, NY 10804.

28-Vdc TO 5-Vdc CONVERTER

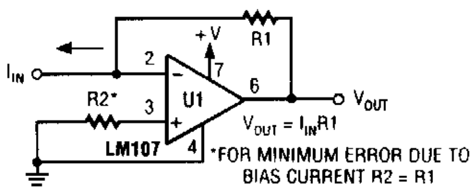


NATIONAL SEMICONDUCTOR

FIG. 18-16

The National Semiconductor LM1575-5.0 allows a very simple switching regulator, with >80% efficiency, operating as a 5-V source @ 1A from a +28-V bus.

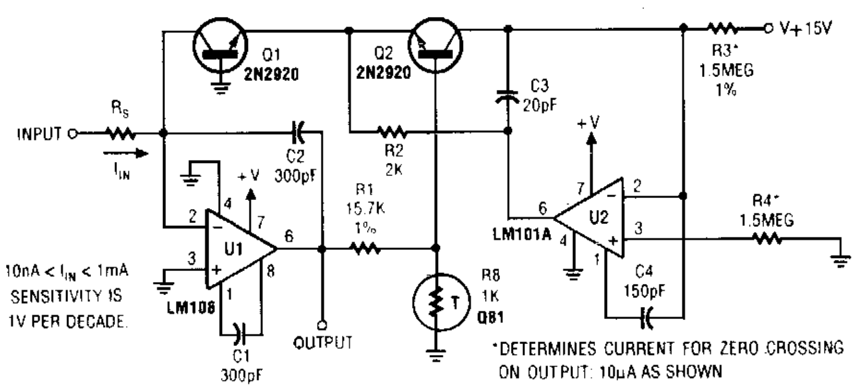
CURRENT-TO-VOLTAGE CONVERTER



POPULAR ELECTRONICS

FIG. 18-17

TEMPERATURE-COMPENSATED ONE-QUADRANT LOGARITHMIC CONVERTER

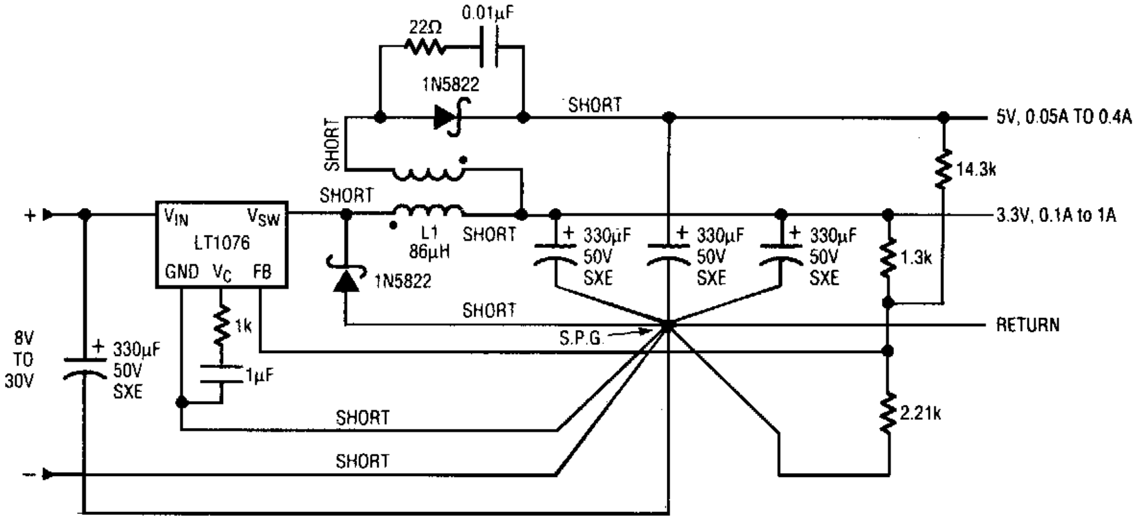


POPULAR ELECTRONICS

FIG. 18-18

A logarithmic converter used to produce an output voltage that is proportional to the logarithm of an input current is shown. R_S is the input impedance of the input source.

dc/dc CONVERTER CIRCUIT WITH 3.3-V AND 5-V OUTPUTS



S.P.G. SINGLE POINT GROUND, (STAR GROUND)
DARK LINES INDICATE HIGH CURRENT PATHS (SEE TEXT)
L1 = HURRICANE LABS HL8685
= COILTRONICS CTX01-11959
ALL ELECTROLYTIC CAPACITORS, UNITED CHEMICON SXE SERIES

LINEAR TECHNOLOGY CORPORATION 1993 ***FIG. 18-19***

Input voltages can range from 8 V to 30 V. The load range on the 5 V is 0.05 A to 5 A while the 3.3-V load range is 0.1 A to 1 A. The circuit is self-protected under no-load conditions. Over all load and line conditions, including cross regulation, the 3.3-V output varies from 3.25 V to 3.27 V. The 5-V output varies from 4.81 V to 5.19 V under the same conditions.

In a typical application to 0.5 A on the 3.3 V and 0.25 A on the 5 V, efficiency is typically 76%. With an input voltage of 30 V and a full-load condition, the efficiency drops to 66%. In normal operating regions, efficiency is always better than 70%. The 5-V ripple is less than 75 mV and the 3.3-V ripple less than 50 mV over all line and load conditions.

19

Counter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

2-MHz FREQUENCY COUNTER

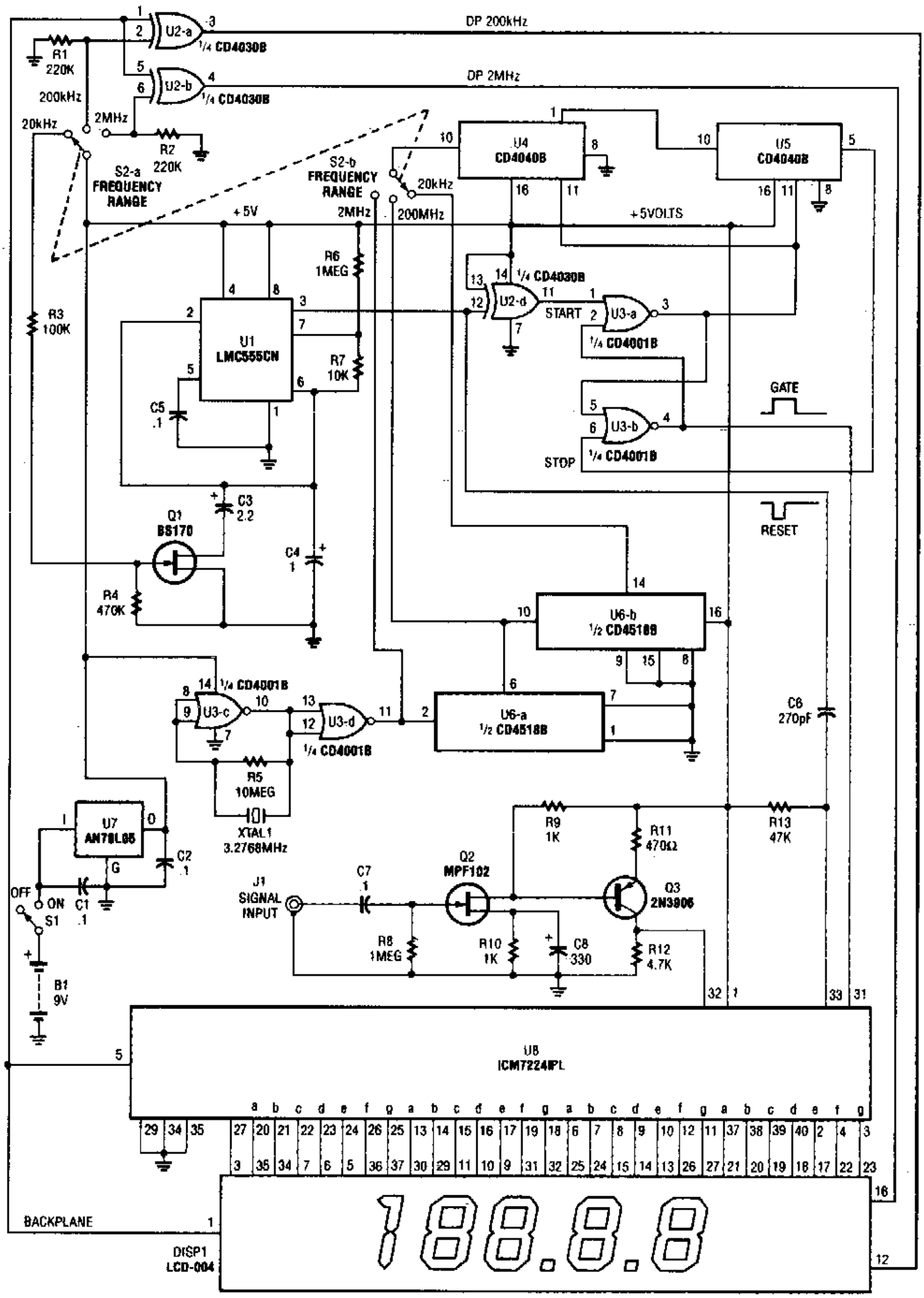
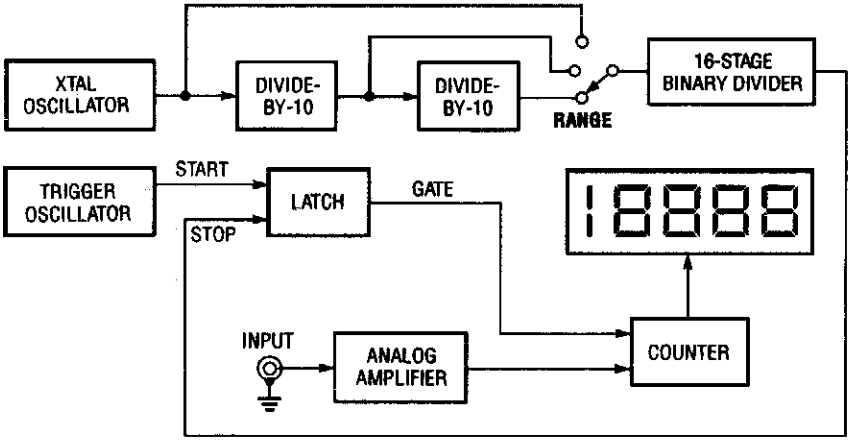


FIG. 19-1



1993 ELECTRONICS HOBBYIST HANDBOOK

This is a schematic and block diagram of a 2-MHz frequency counter. It uses and LSI counter/display driver, LCD readout, and a few logic chips for timebase and timing pulse circuitry. Q2 and Q3 form a signal (input) amplifier.

The circuit contains a crystal oscillator built around U3-c and XTAL1, which provides the primary timing-reference signal. That signal is then divided twice to provide two additional timing references, giving the circuitry three selectable timing references. The ICM7224IPL is an integrated circuit that consists of the counter and display driver to drive the LCD-004 display.

10-MHz FREQUENCY COUNTER

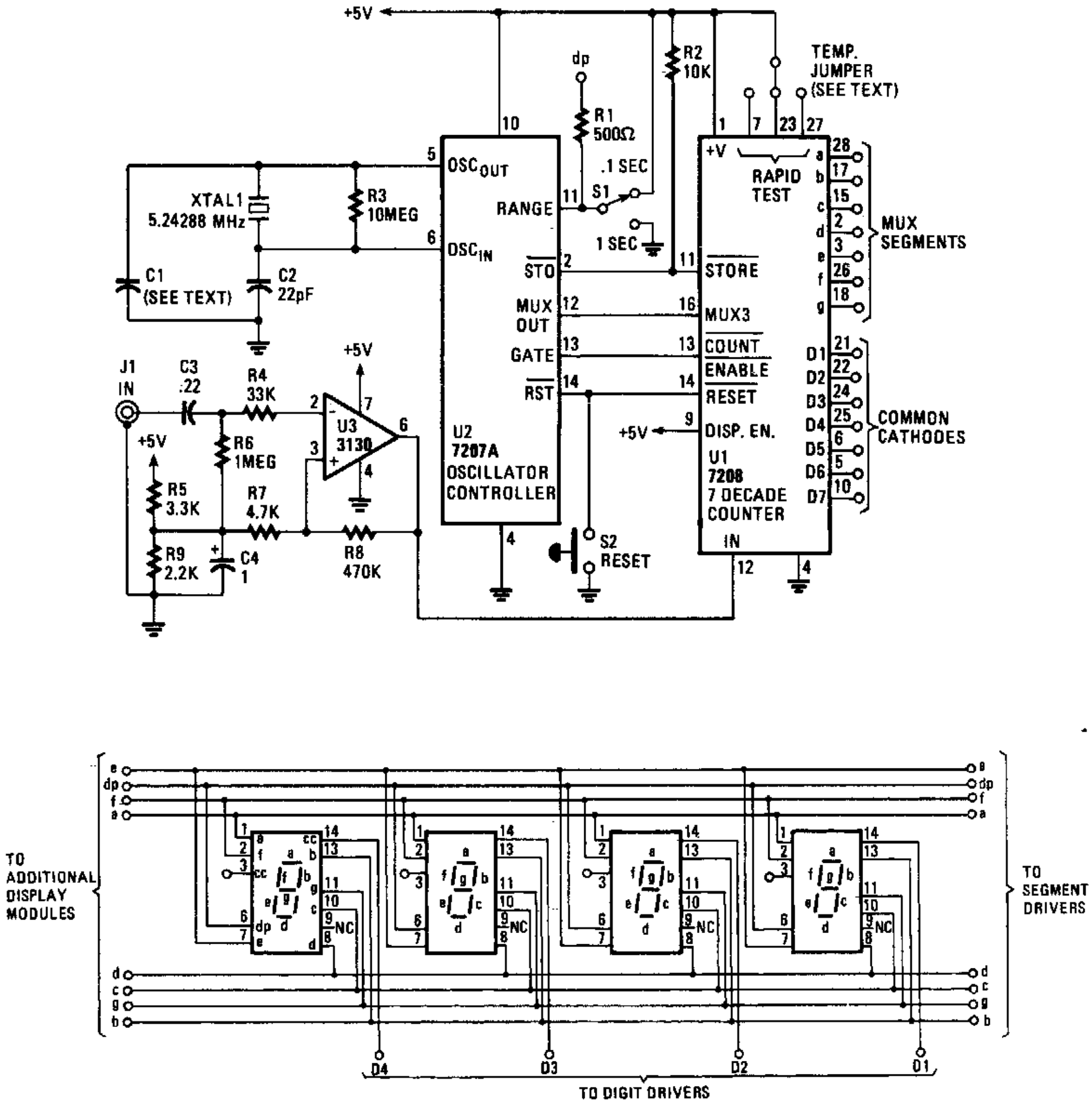


FIG. 19-2

The circuit consists of an ICM7208 seven-decade counter (U1), an ICM7207A oscillator controller (U2), and a CA3130 biFET op amp (U3). Integrated circuit UI counts input signals, decodes them to 7-segment format, and outputs signals that are used to drive a 7-digit display. Integrated circuit U2 provides the timing for UI, while U3 conditions the input signal to provide a suitable waveform for input to U1. The 5.24288-MHz crystal frequency is divided by U2 to produce a 1280-Hz multiplexing signal at pin 12 of U2. That signal is input to U1 at pin 16 and is used to scan the display digits in sequence. The cathodes of each digit are taken to ground several times each second, activating any segments of the digits whose anodes are high as the result of decoding by U1. The crystal frequency is further divided to produce a short "store" pulse at pin 2 of U2, followed (after about 0.4 ms) by a short "reset" pulse at pin 14 of U2. The frequency of the pulses is determined by the state of U2 pin 11.

When pin 11 of U2 is taken to ground through S1, the pulses occur every 2 seconds and cause U2 pin to go high for one second, which prevents additional input signals from entering Ul. That causes the count latched in U1's internal counters to be transferred to the display.

Integrated circuit U2 pin 13 then goes low for one second, allowing a new count to be entered into the seven decade counters of U1. That cycle is repeated, continuously updating the display every two seconds.

When U2 pin 11 is taken to the positive supply rail (+5 V), the "store" and "reset" pulses occur at 0.2-s intervals, resulting in a 0.1-s count-period. Ten input pulses must be counted in order for a "1" to appear on the first digit, D1, so that the frequency being measured is obviously 10 times larger than the frequency that is shown on the display. In that mode, the decimal points are driven by M and visually indicate that the 0.1-s count period is being used.

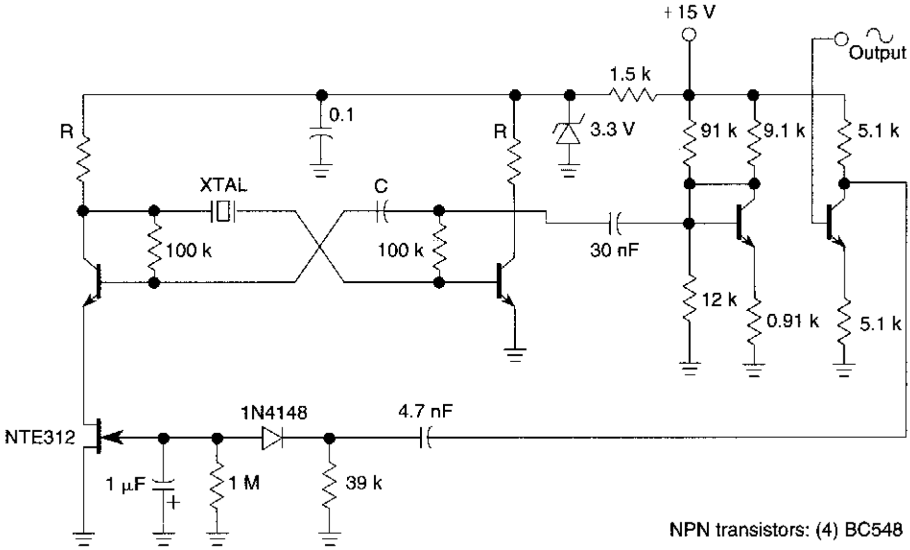
The display must have at least seven 7-segment common-cathode multiplexed LED digits. Any common-cathode seven-segment display can be used; no particular display is specified.

20

Crystal Oscillator and Test Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

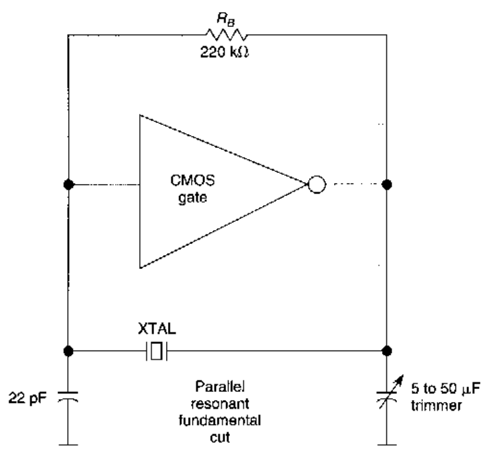
LOW-FREQUENCY CRYSTAL OSCILLATOR



RF DESIGN ***FIG. 20-1***

Q1, Q2, and the associated circuitry form a modified astable multivibrator in which the loop gain is automatically adjusted to the threshold of oscillation by means of field effect transistor Q3. Q4 linearly amplifies the signal present at the collector of Q2 and isolates the oscillator section of the circuit from the output. This stage features wideband operation and delivers a clean 2.5-V amplitude sine wave into a resistive load greater than or equal to 20 kΩ. The stage comprising Q5 has a voltage gain of 1 and its sole purpose is to isolate the nonlinear effects of rectifier D1 from the output.

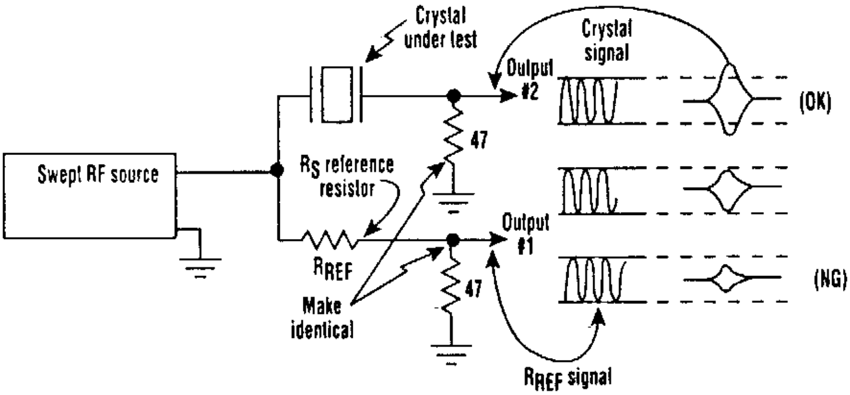
CRYSTAL OSCILLATOR



The CMOS amplifier is biased into the linear region by resistor R_B. The pi-type crystal network (C1 and C2, and XTAL) provides the 180° phase shift at the resonant frequency which causes the circuit to oscillate.

WILLIAM SHEETS ***FIG. 20-2***

EASY CRYSTAL IMPEDANCE CHECKER



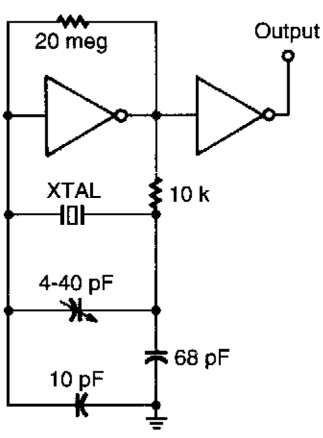
ELECTRONIC DESIGN

FIG. 20-3

On occasion, microprocessors/microcomputers and microprocessor crystals just aren't compatible with each other. Many microprocessor data sheets specify maximum values for a crystal's equivalent series resistance (R_S) that aren't met by some crystals advertised for microprocessor/microcomputer use. As a result, a crystal with an R_S value greater than the maximum specified for the chip might cause problems, such as a balky or even inoperative clock oscillator.

To tackle this problem, a suspected crystal can be given a quick check for R_S with a simple test setup that consists of a sweep generator, oscilloscope, and three resistors (see the figure). When the frequency source is brought to the crystal's frequency, output 2 will maximize. If it exceeds the amplitude of output 1, the crystal's R_S value will be less than the R_S reference resistor's value. If it doesn't exceed output 1's amplitude, the crystal's R_S value is too large.

HEX BUFFER CRYSTAL OSCILLATOR

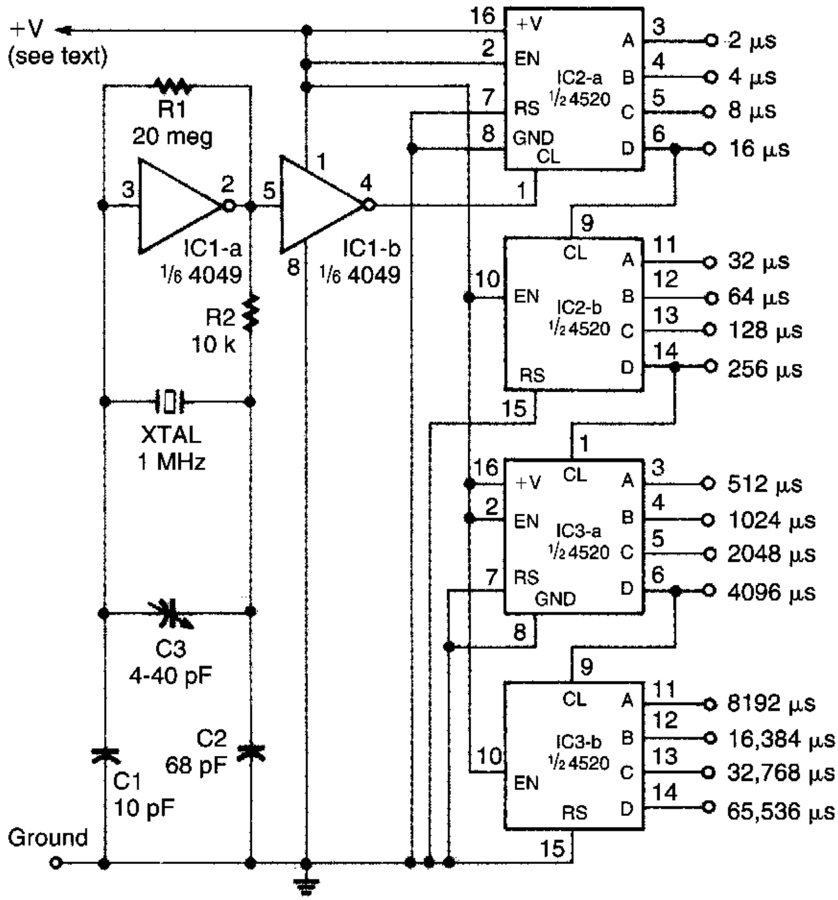


A 4049 single section acts as a crystal oscillator, driving another section as a buffer, leaving four sections for other use. Use a 32- or 20-pF parallel resonant fundamental crystal.

ELECTRONICS NOW

FIG. 20-4

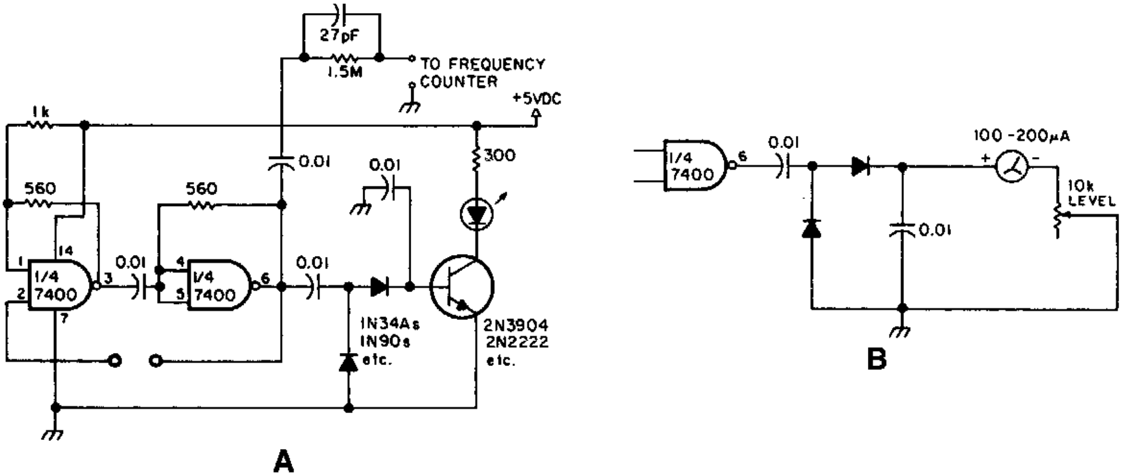
MULTI-OUTPUT TIMEBASE



RADIO-ELECTRONICS *FIG. 20-5*

A 1-MHz oscillator drives a binary counter to produce pulse widths from 2 to 65,536 ms. *V*+ is any CMOS suitable level (5 to 15 V, etc.).

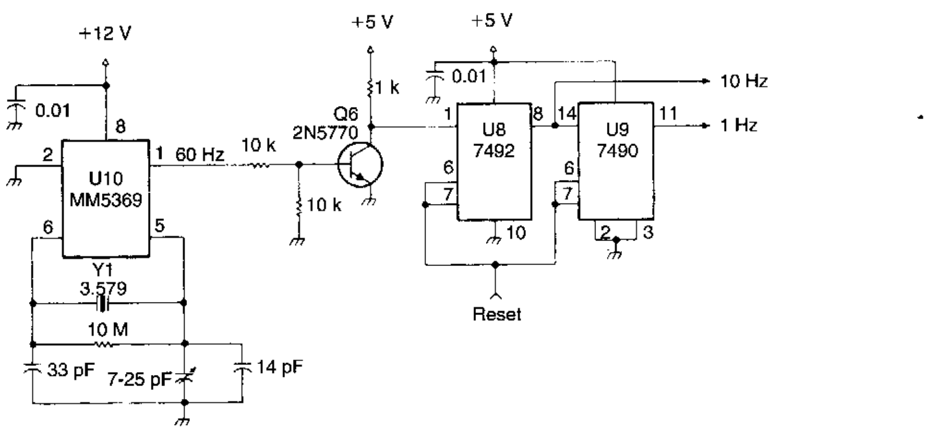
CRYSTAL ACTIVITY TESTER



73 AMATEUR RADIO TODAY *FIG. 20-6*

This circuit will check a crystal for activity. Two sections of a 7400 act as an oscillator and its output is rectified and drives an npn transistor that switches an LED (Fig. A). In Fig. B, a meter replaces the LED.

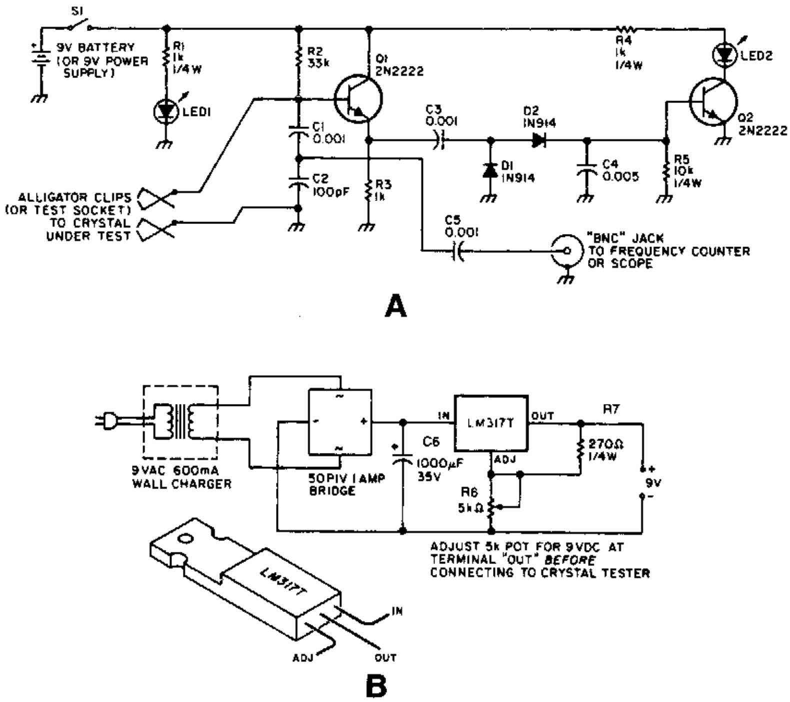
10- TO 1-Hz TIMEBASE



73 AMATEUR RADIO TODAY *FIG. 20-7*

This system uses an MM5369 IC to derive a 60-Hz signal from a TV burst crystal (3579 MHz). *V8* and V9 produce a 10-Hz and 1-Hz signal from this 60-Hz signal. Y1 can be any parallel-mode 3.579-MHz crystal.

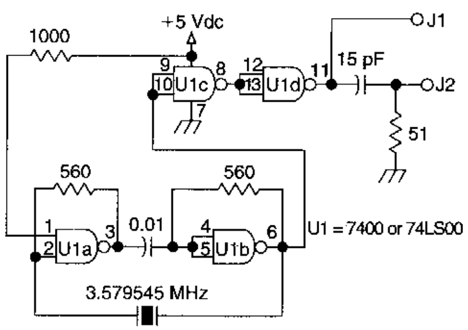
CRYSTAL TESTER



73 AMATEUR RADIO TODAY ***FIG. 20-8***

Q1 acts as a Colpitts crystal oscillator, and if the crystal under test is operational, the RF signal is rectified by D1 and D2, turning on Q2 and lighting indicator LED2. LED1 is a power indicator.

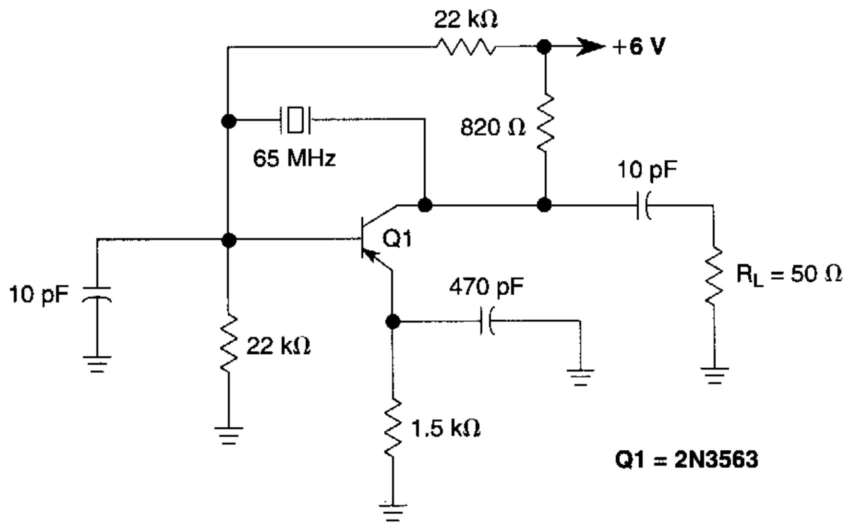
WIDE-RANGE CRYSTAL OSCILLATOR



A circuit using one 7400 TTL IC can use crystals of the fundamental type, from 1 to about 13 MHz. Output is rich in harmonics, making this oscillator useful for calibrations and test applications.

73 AMATEUR RADIO TODAY ***FIG. 20-9***

PIERCE OSCILLATOR

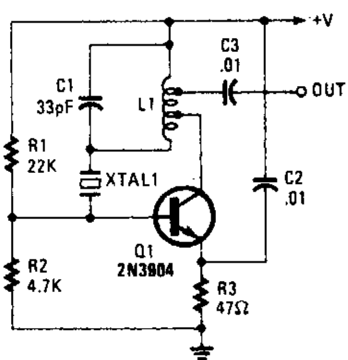


WILLIAM SHEETS

FIG. 20-10

This Pierce oscillator uses a fundamental-mode 65-MHz crystal.

CRYSTAL-CONTROLLED HARTLEY OSCILLATOR



POPULAR ELECTRONICS

FIG. 20-11

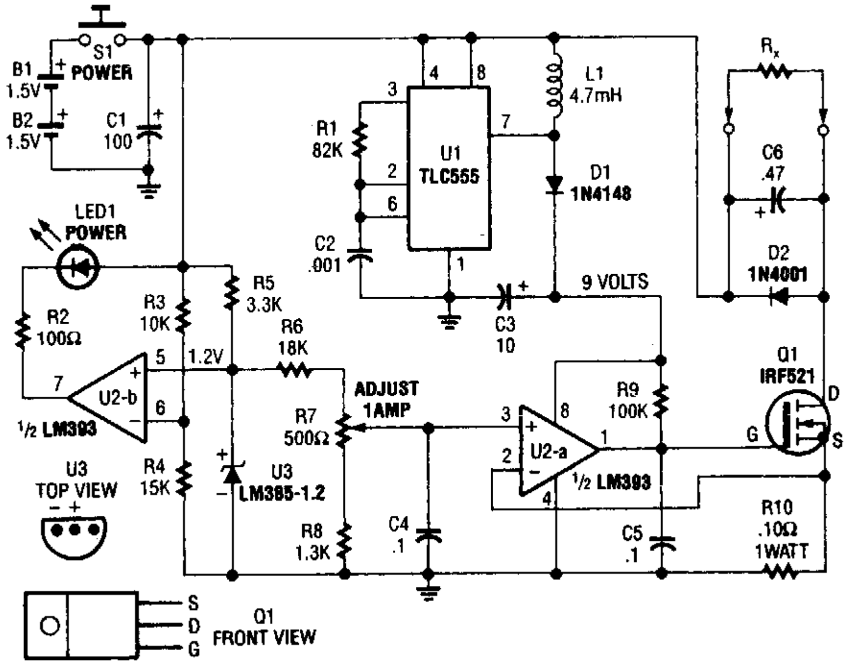
21

Current-Source Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Current Source for Low-Resistance Measurements
Precision Positive Current Source
Bilateral Current Source
Precision Negative Current Source

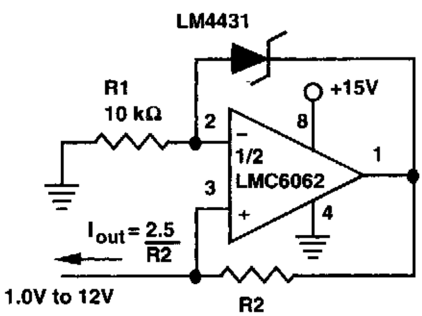
CURRENT SOURCE FOR LOW-RESISTANCE MEASUREMENTS



1993 ELECTRONICS HOBBYISTS HANDBOOK *FIG. 21-1*

Useful for low-resistance measurements, this 1-A current source will produce 1 A in unknown resistance R_x. For best results, R_x should be less than 1 to 2 Ω, because only 3 V are available. U1 is a flyback converter to generate 9 V for U2.

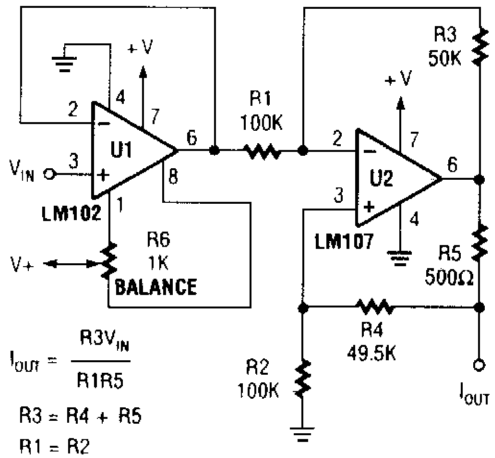
PRECISION POSITIVE CURRENT SOURCE



An LM4431 precision 2.5-V reference and an LMC6062 op amp to make a positive current source, from 1 mA to 10 mA.

NATIONAL SEMICONDUCTOR *FIG. 21-2*

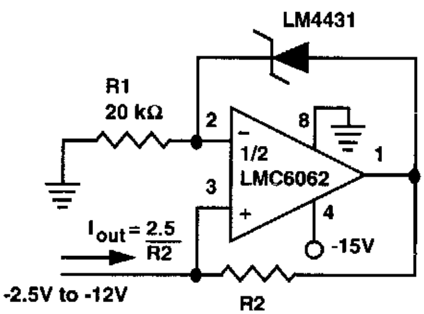
BILATERAL CURRENT SOURCE



POPULAR ELECTRONICS *FIG. 21-3*

Using two op amps, this circuit produces current proportional to V_{IN}.

PRECISION NEGATIVE CURRENT SOURCE



NATIONAL SEMICONDUCTOR *FIG. 21-4*

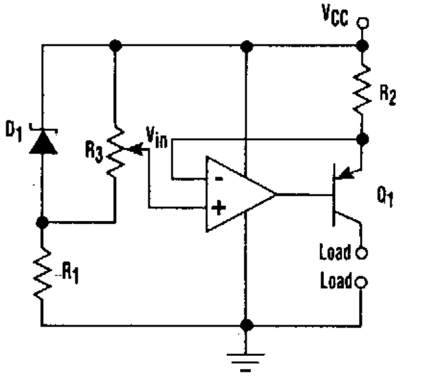
A National Semiconductor LM4431 reference and an LMC6062 op amp make up a negative current source. Current range is 1 μA to 1 mA.

22

Current Limiter and Control Circuits

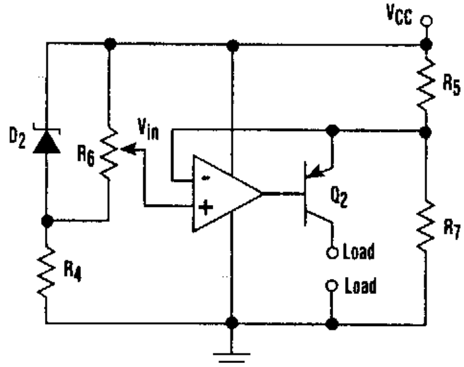
The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

OFFSET-ADJUSTING CURRENT SOURCE



1. Though this setup can act as a cost-effective current source with an output accurate to 1%, the voltage offset will turn on the current source even when V_{CC} equals V_{in}.

A



2. Modifying the configuration of FIgure 1 can rectify the problem of the current source being turned on by the voltage offset. The addition of R_7 allows an adjustment that guarantees turn-off for any op-amp offset specification.

B

FIG. 22-1

By carefully choosing components, you can create a cost effective circuit for a current source with an output that's accurate to 1% (Fig. A). I_{OUT} (the current flowing from the collector of Q1) is V_{CC}–V_{IN} (the voltage at the wiper of R3) divided by the value of R_2.

In some instances, it's important to be able to turn off the current source (within the limits of I_{CEO} for Q1). Unfortunately, in about half of these cases, the offset voltage (V_{OS}) of the op amp will turn the current source on even when $V_{CC} = V_{IN}$. That's because the offset voltage (when the noninverting input needs to be at a higher potential than the inverting input to get an output of 0 V from the op amp) is impressed across R2. This offset voltage forces Q1 to turn on enough to yield a collector current of V_{OS} divided by R_2.

Figure B offers a fix for this predicament. The addition of R7 presents the emitter of Q2 with a Thevenin equivalent voltage and resistance represented by:

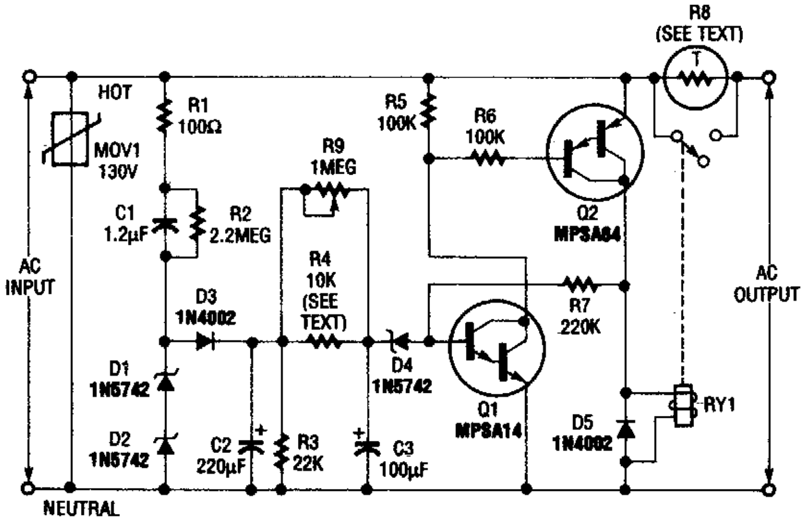
$$V_{TH} = \frac{V_{CC}\,(1 - R_5)}{R_5 + R_7}$$

$$R_{TH} = \frac{R_5 \times R_7}{R_5 + R_7}$$

The difference between V_{CC} and V_{TH} is $V_{CC}\,(R_5/R_5 + R_7)$. If $V_{CC}\,(R_5/R_5 + R_7)$ is set equal to the maximum V_{OS} spec for the op amp in question, the circuit is then guaranteed to turn off. This circuit has an output current of $V_{TH} - V_{IN}$ divided by R_{TH}.

The compromise of Fig. B does present another error term in the circuit. The term $(V_{TH} - V_{IN})$ will have to be $2 \times V_{OS}$ to guarantee a current output for whole population of the op amp chosen. This error can be made arbitrarily small (but not zero) by increasing the voltage of D2 and V_{CC} while raising the value of D2 and V_{CC} while also raising the value of the equivalent resistance R_{TH}.

INRUSH CURRENT LIMITER



ELECTRONICS NOW

FIG. 22-2

Q1 is an npn Darlington and Q2 is a pnp Darlington. MOV1 is a metal-oxide varistor and R8 is an NTC thermistor for limiting inrush current.

This circuit limits ac line current to a load. When a predetermined interval has passed, RY1 shorts out thermistor or resistance *RB*. R4 can be 150 kΩ if R9 is not used. If power is removed, the circuit is ready for immediate restart.

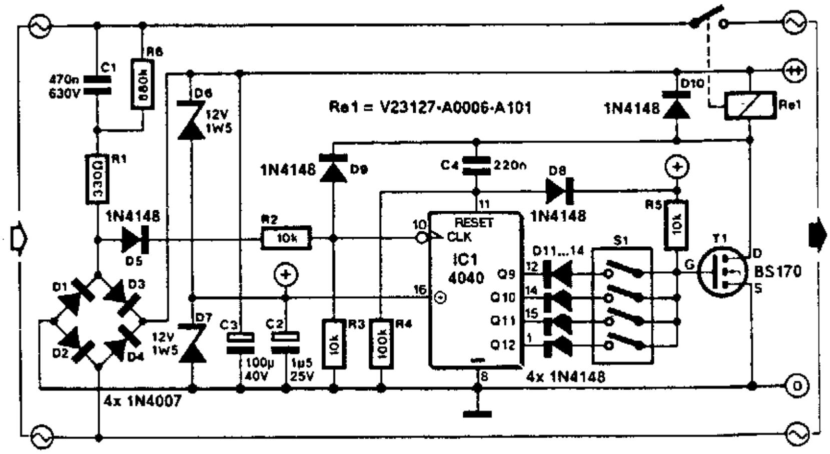
23

Delay Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Power-On Delay Circuit

POWER-ON DELAY CIRCUIT



DELAY TIMES				
Switches S_1				Time (s)
1	2	3	4	
0	0	0	0	0.0
0	0	0	1	5.1
0	0	1	0	10.2
0	0	1	1	15.4
0	1	0	0	20.5
0	1	0	1	25.6
0	1	1	0	30.7
0	1	1	1	35.8
1	0	0	0	41.0
1	0	0	1	46.1
1	0	1	0	51.2
1	0	1	1	56.3
1	1	0	0	61.4
1	1	0	1	66.6
1	1	1	0	71.7
1	1	1	1	76.8

ELEKTOR ELECTRONICS ***FIG. 23-1***

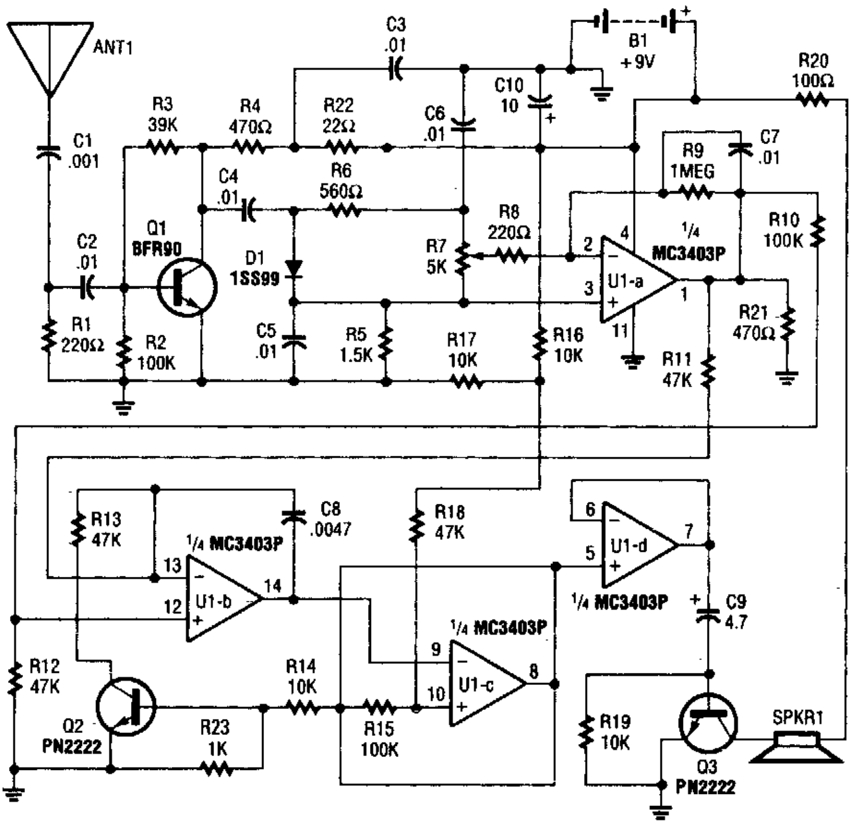
Using an IC to count ac mains pulses, the circuit produces 16 various delay times, where ac power is applied to a load after a preset interval.

24

Detector, Demodulator, and Discriminator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

BUG DETECTOR



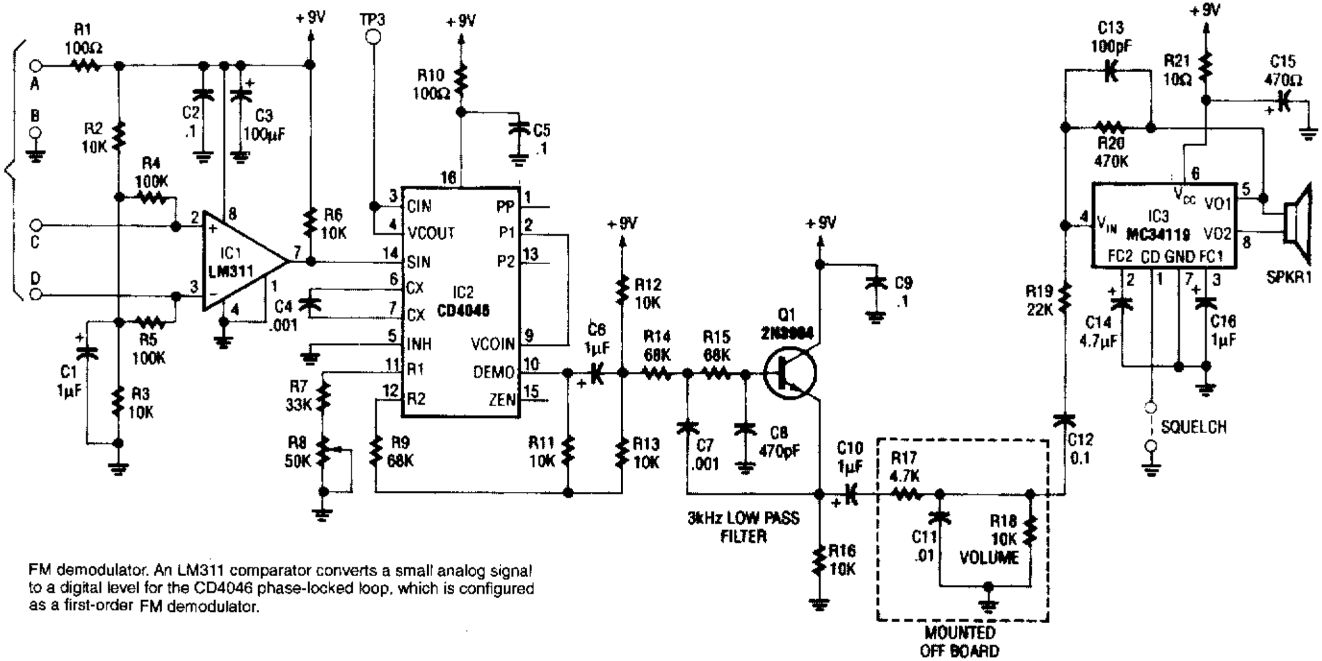
1992 PE HOBBYIST HANDBOOK ***FIG. 24-1***

The circuit, built around a single integrated circuit (U1, an MC3403P quad op amp), three transistors (Q1–Q3), and a few support components, receives its input from the antenna (ANT1). The signal is fed through a high-pass filter, formed by C1, C2, and R1, which eliminates bothersome 60-Hz pickup from any nearby power lines or line cords located in and around buildings and homes.

From the high-pass filter, the signal is applied to transistor Q1 (which provides a 10-dB gain for frequencies in the 1- to 2000-MHz range) for amplification. Resistors R2, R3, and R4 form the biasing network for Q1. The amplified signal is then ac coupled, via capacitor C4 and resistor R7's (the sensitivity control) wiper, to the inverting input (pin 2) of U1-a. Op amp U1-a is configured as a very high gain amplifier. With no signal input from ANT1, the output of U1-a at pin 1 is near ground potential.

When a signal from the antenna is applied to the base of Q1, it turns on, producing a negative-going voltage at the cathode of D1. That voltage is applied to pin 1 of U1-a, which amplifies and inverts the signal, producing a positive-going output at pin 1. Op amps U1-b and U1-c along with C8, R10 through R18, and Q2 are arranged to form a voltage-controlled oscillator (VCO) that operates over the audio-frequency range. As the output of U1-a increases, the frequency of the VCO increases. The VCO output, at pin 8 of U1-c, is fed to the input of U1-d, which is configured as a noninverting, unity-gain (buffer) amplifier. The output of U1-d is used to drive Q3, which, in turn, drives the output speaker.

FM DEMODULATOR



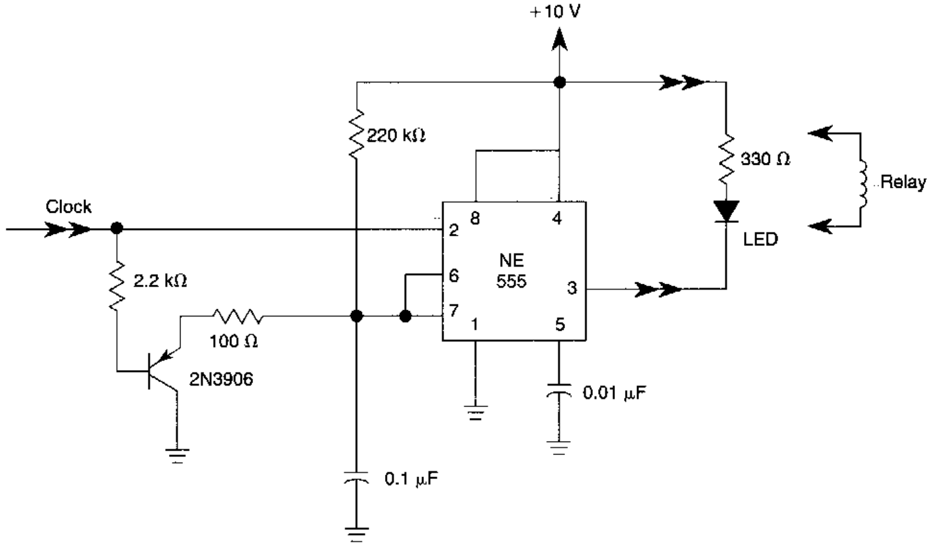
FM demodulator. An LM311 comparator converts a small analog signal to a digital level for the CD4046 phase-locked loop, which is configured as a first-order FM demodulator.

ELECTRONICS NOW

FIG. 24-2

An LM311 comparator converts a small analog signal to a digital level for the DC4046 phase-locked loop, which is configured as a first-order FM demodulator. This demodulator works with a 50-kHz FM modulated input signal. It has applications in FM light beam receivers or in remote control applications. Pin 1 of IC3 can be used to squelch the receiver if it is lifted from ground; if not desired, leave it grounded.

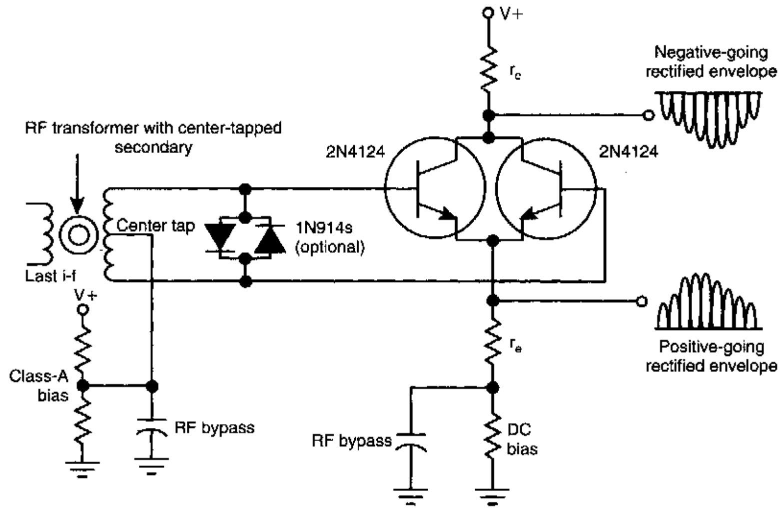
555 MISSING PULSE DETECTOR



WILLIAM SHEETS ***FIG. 24-3***

This missing pulse detector can use an LED or relay output.

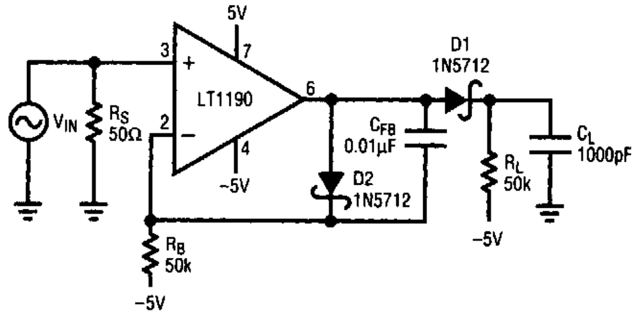
SIMPLE FULL-WAVE ENVELOPE DETECTOR



ELECTRONIC DESIGN ***FIG. 24-4***

Simple, yet sensitive, this amplifying full-wave detector circuit has an almost zero rectification threshold. It presents a highly linear RF load to the final IF stage. The gain for the collector output is given (approximately) by r_c/r_e. The emitter output gain is slightly less than unity.

OPEN-LOOP PEAK DETECTOR

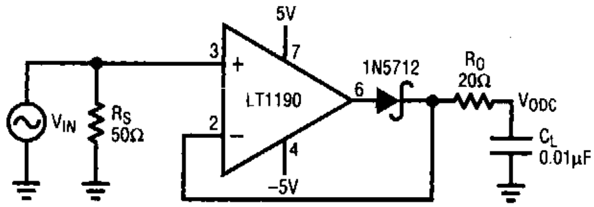


LINEAR TECHNOLOGY

FIG. 24-5

In this open-loop design, the detector diode is D1, and a level shifting or compensating diode is D2. Load resistor R_L is connected to –5 V, and an identical bias resistor R_L is connected to –5 V, and identical bias resistor R_B is used to bias the compensating diode. Resistors with equal values ensure that the diode drops are equal. Low values of R_L and R_B (1 kΩ to 10 kΩ) provide fast response, but at the expense of poor low-frequency accuracy. High values of R_L and R_B provide good low-frequency accuracy, but cause the amplifier to slew rate limit, resulting in poor high-frequency accuracy. A good compromise can be made by adding a feedback capacitor C_{FB}, which enhances the negative slew rate on the (–) input.

CLOSED-LOOP PEAK DETECTOR

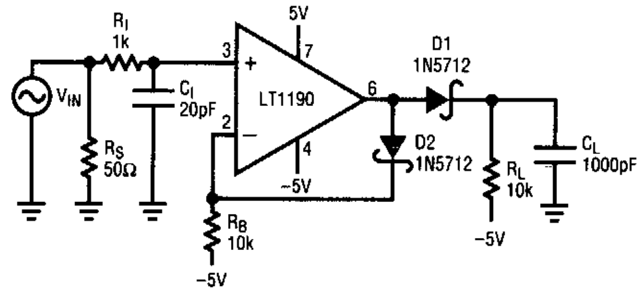


LINEAR TECHNOLOGY

FIG. 24-6

This closed-loop peak detector circuit uses a Schottky diode inside feedback loop to obtain good accuracy. The 20-Ω resistance R_O isolates the 0.01-µF load and prevents oscillation. The dc value is read with a DVM. At a low frequency, the error is small and dominated by the decay of the detector capacitor between cycles. As the frequency rises, the error increases because capacitor charging time decreases. During this time, the overdrive becomes a very small portion of a sine-wave cycle. Finally, at approximately 4 MHz, the error rises rapidly because of the slew-rate limitation of the op amp.

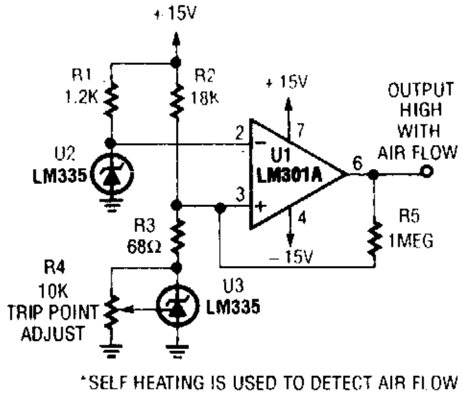
FAST PULSE DETECTOR



LINEAR TECHNOLOGY *FIG. 24-7*

A fast pulse detector can be made with this circuit. A very fast input pulse will exceed the amplifier slew rate and cause a long overload recovery time. Some amount of dv/dt limiting on the input can help this overload condition, however this will delay the response.

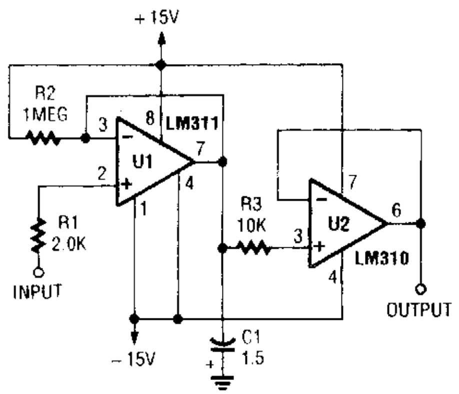
AIR-FLOW DETECTOR



POPULAR ELECTRONICS *FIG. 24-8*

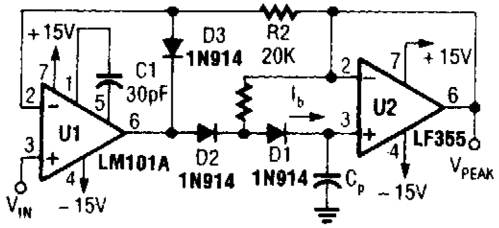
Two precision temperature sensors are used to detect a small temperature difference. When air flow occurs, self-heating of the LM335 is reduced, and the output of the two temperature sensors is unequal. This is amplified by U1.

NEGATIVE PEAK DETECTOR



POPULAR ELECTRONICS *FIG. 24-9*

LOW-DRIFT PEAK DETECTOR



Leakage of D2 is provided by feedback path through R_f.

Leakage of circuit is essentially I_b (LF155, LF156) plus capacitor leakage of Cp.

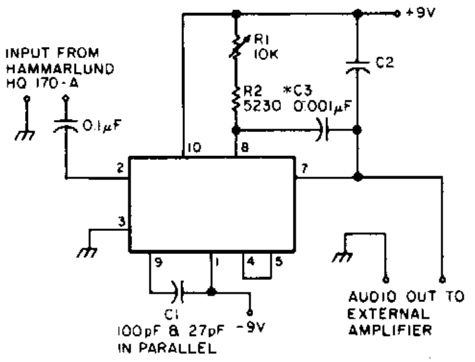
Diode D3 clamps V_{out} (A1) to $V_{in} - V_{D3}$ to improve speed and to limit reverse bias of D2.

Maximum input frequency should be <<½π R_fC_{D2}, where C_{D2} is shunt capacitance of D2.

POPULAR ELECTRONICS

FIG. 24-10

455-kHz FM DEMODULATOR



*C3 IS REQUIRED TO ELIMINATE POSSIBLE OSCILLATION IN THE CONTROL CURRENT SOURCE

73 AMATEUR RADIO

FIG. 24-11

Free-running frequency of VCO: $f_{os} = 1.2/4\,(R_1)\,(C_1)$

$$\text{lock range } f_1 = \pm 8 f_{os}/V_{CC}$$

$$\text{capture range } f_c = \pm \tfrac{1}{2}\pi \sqrt{\frac{2\pi F_L}{r}}$$

$$\text{where } r = (3.6 \times 10^3)\,(C2)$$

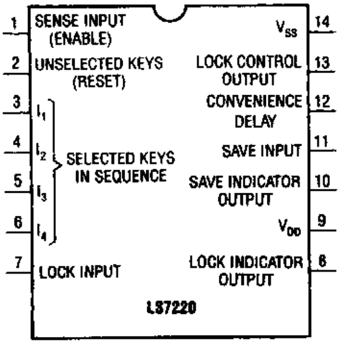
Useful for NBFM reception on older shortwave receivers lacking this capability, this circuit uses a PLL IC, an N565N, to achieve this. It was originally used with an old Hammarlund HQ-170 receiver, for both 6- and 10-m FM reception.

25

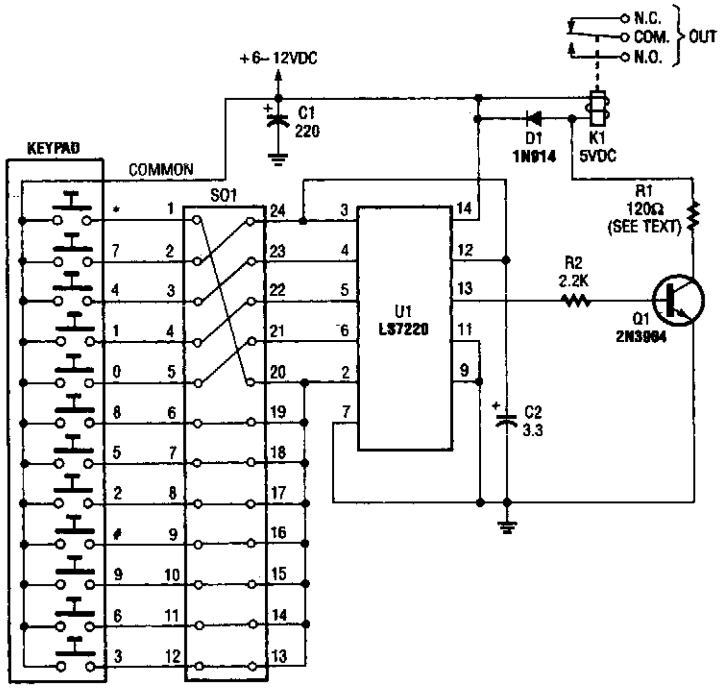
Digital Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Digital Entry Lock
Digital Audio Selector
Digital Multiple-Gang Potentiometer Control
Digital Resistance Control
Digital Capacitance Control
BCD Rotary Switch



The LS7220 keyless lock (a pinout of which is shown here) is a special-purpose IC designed to accept a four-digit code.



1992 PE HOBBYIST HANDBOOK ***FIG. 25-1***

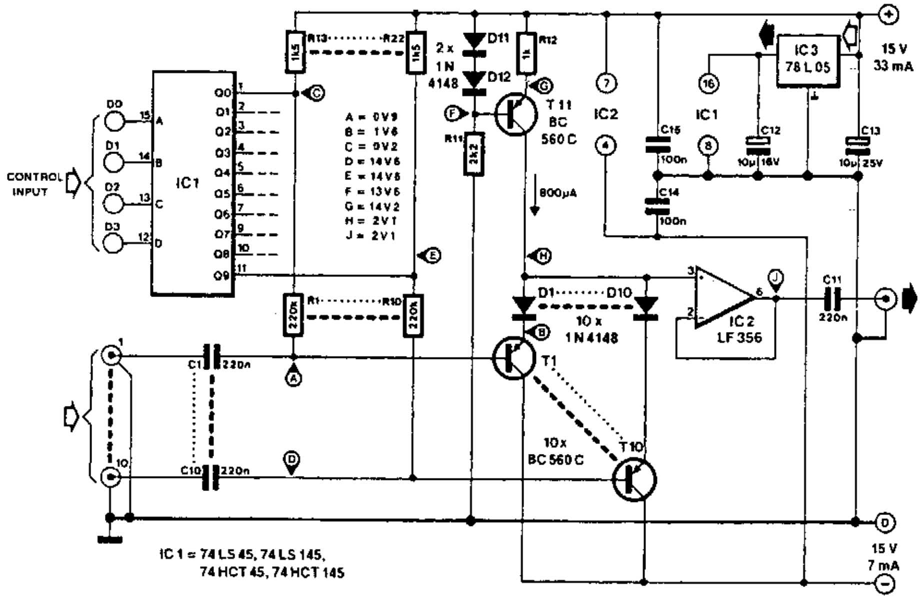
A block pinout diagram of the LS7220 keyless-lock IC is shown. The keypad must provide each key with a contact to a common connection. In this case, the common connection goes to the positive supply rail so that when a key is pressed, a positive voltage is passed through to the wire associated with that key. Each of the 12 keys are brought out to separate wires, and each wire is connected to a different pin of a 24-pin socket (SO1).

To activate (unlock) the circuit, a preprogrammed four-digit access code must be entered in the proper sequence. The four-digit access code must be entered in the proper sequence. The four-digit access is programmed into the circuit by connecting jumpers between terminals of a 24-pin plug-in header.

When the correct access code is entered (in the proper sequence), positive voltages appear at pins 3, 4, 5, and 6 of U1. That causes U1 to output a positive voltage at pin 13, which is fed through resistor R2 to the base of Q1, causing it to conduct. With Q1 conducting, its collector is pulled to ground potential, energizing relay K1. The normally open relay contacts close, switching on any external device.

Capacitor C2 controls the total time that the output of U1 at pin 13 is positive after the release of the first key. With a value of 3.3 μF for C2, active time after release of the first key is about two seconds, assuming a 6-V supply or four seconds with a 12-V supply. Therefore, if you push the subsequent keys too slowly, the relay might not close at all! To increase the time allotted for code entry, you will have to increase the capacitance of C2.

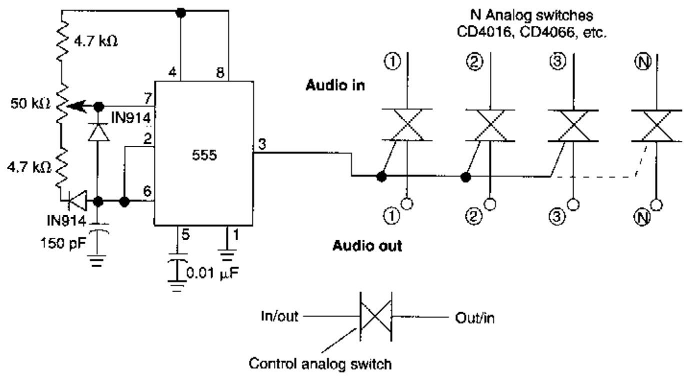
DIGITAL AUDIO SELECTOR



303 CIRCUITS *FIG. 25-2*

This circuit uses switched emitter followers, rather than the usual analog switch CMOS chips. This yields better reduction of crosstalk between channels. This circuit can handle up to 4 V_{rms} with less than −80-dB crosstalk.

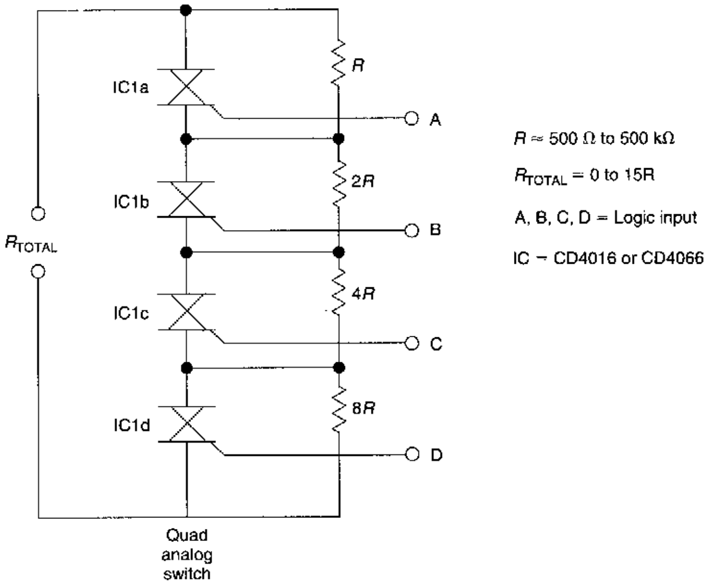
DIGITAL MULTIPLE-GANG POTENTIOMETER CONTROL



WILLIAM SHEETS *FIG. 25-3*

A 555 timer can be configured to simulate a multi-gang potentiometer by controlling the mark-space ratio. The switching rate should be at least twice the maximum expected signal frequency the potentiometer has to handle.

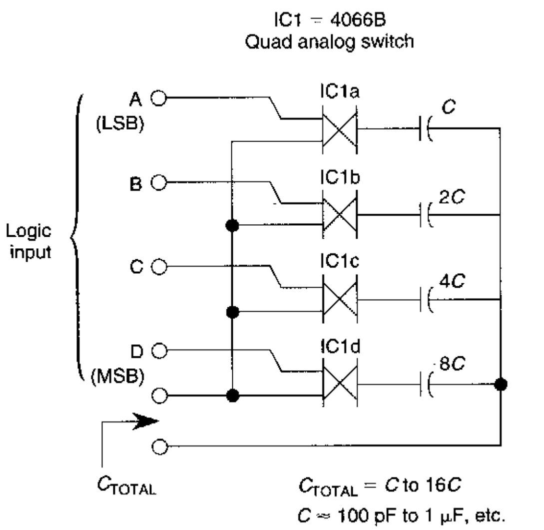
DIGITAL RESISTANCE CONTROL



WILLIAM SHEETS **FIG. 25-4**

Digital resistance control is possible with bilateral switches. Do not forget that analog switches have "on" resistance.

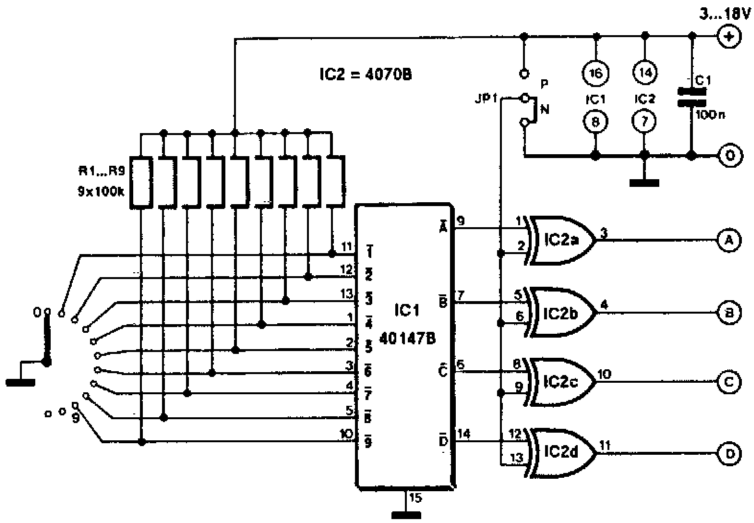
DIGITAL CAPACITANCE CONTROL



Digital capacitance control is possible with bilateral switches. Do not forget to consider "ON" resistance of the analog switches.

WILLIAM SHEETS **FIG. 25-5**

BCD ROTARY SWITCH



ELEKTOR ELECTRONICS USA

FIG. 25-6

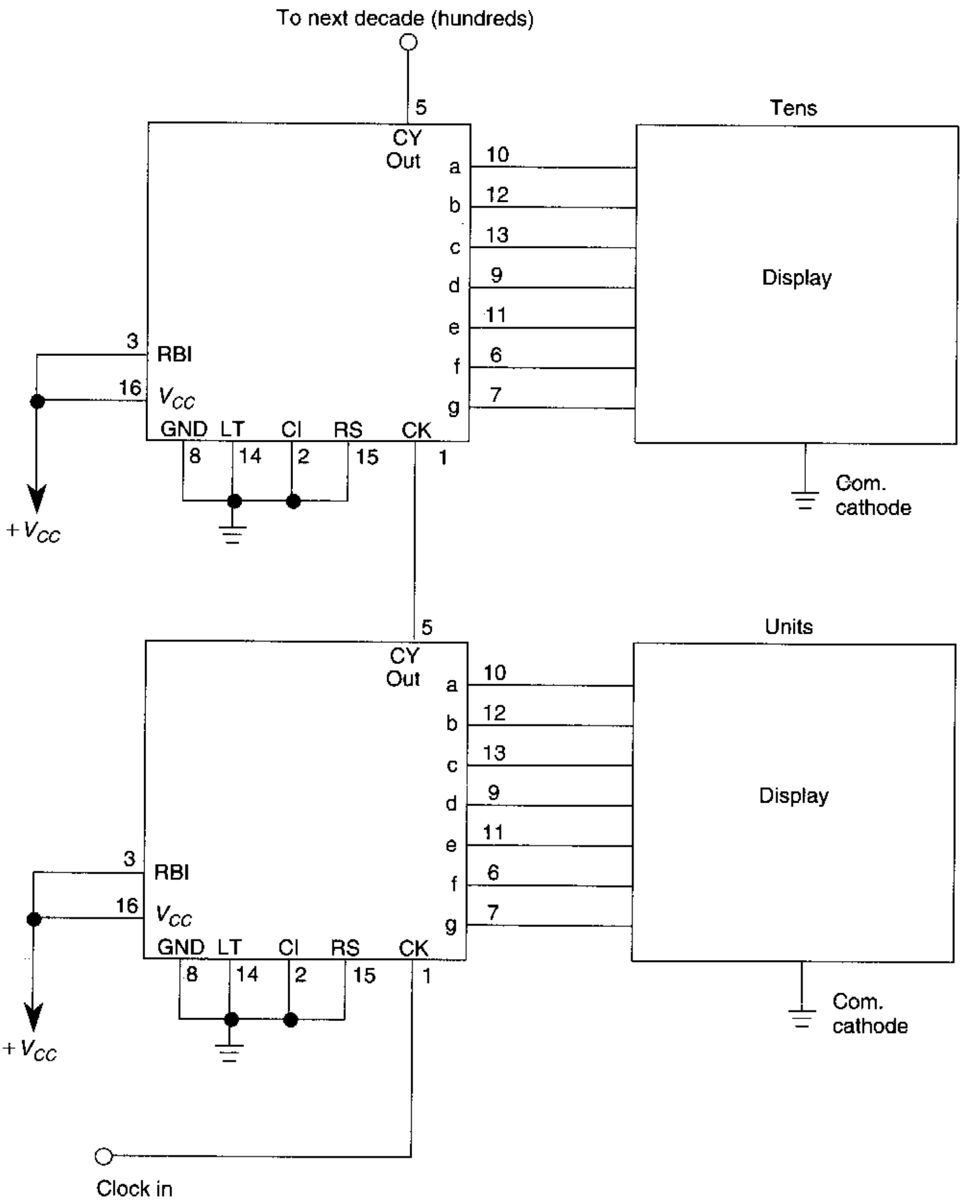
This circuit allows a simple rotary switch to emulate a BCD switch. The circuit draws about 200 mA. A 10-position rotary switch is used.

26

Display Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

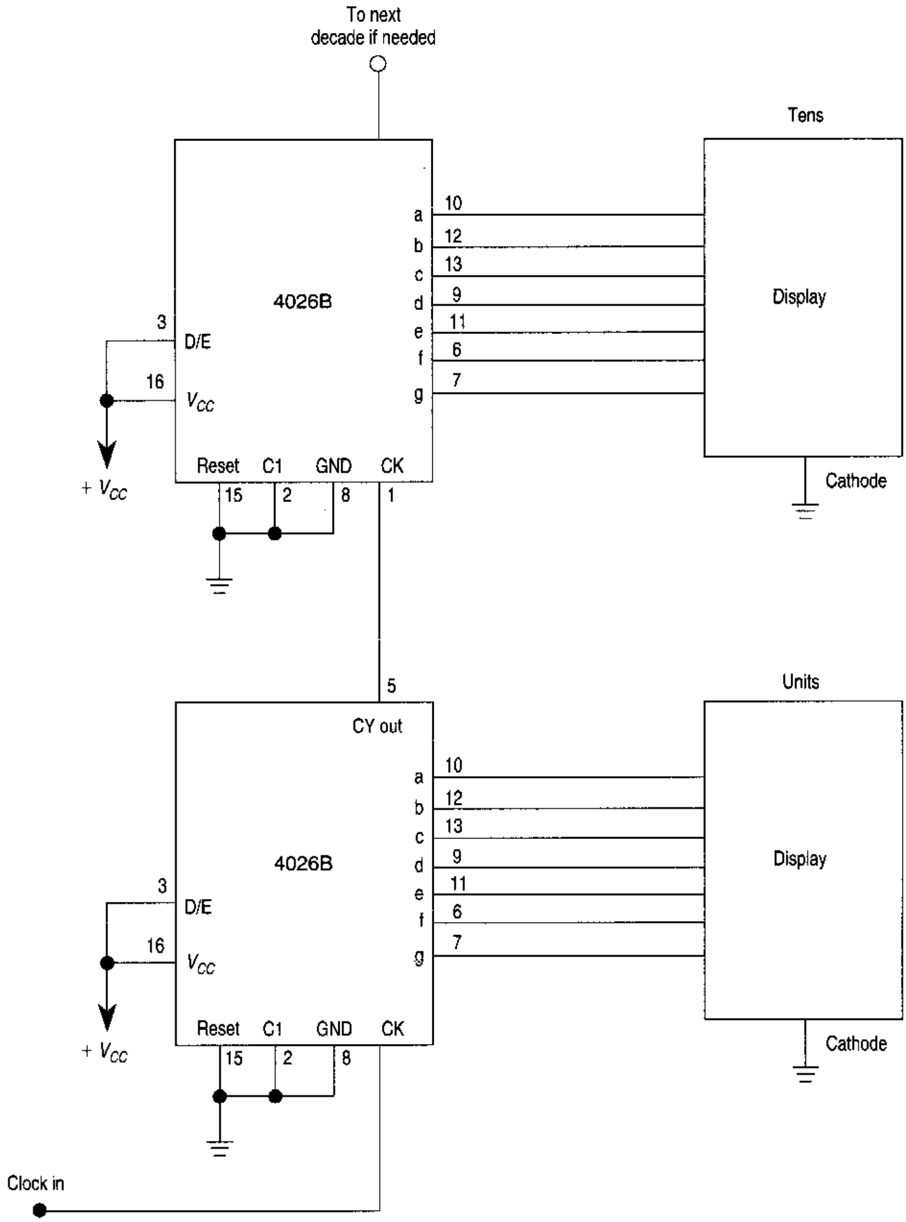
4033 DISPLAY CIRCUITRY COMMON CATHODE



WILLIAM SHEETS **FIG. 26-1**

To drive two or more common-cathode displays two or more 4033 decode counters can be cascaded.

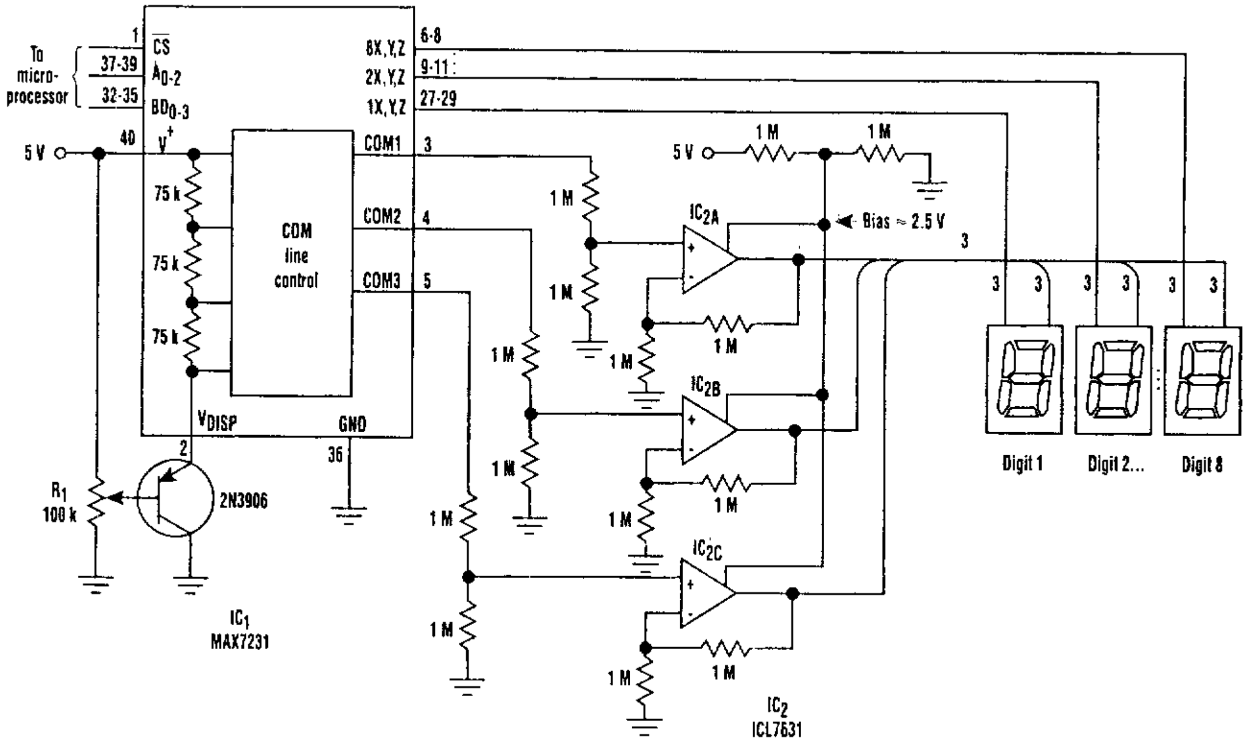
CASCADED 4026B COUNTER/DISPLAY DRIVER CIRCUIT



WILLIAM SHEETS ***FIG. 26-2***

Two or more 4026B counters can be cascaded as shown to give a multiple-digit display. Two, three or more displays can thus be connected.

LARGE LCD DISPLAY BUFFERING DRIVER



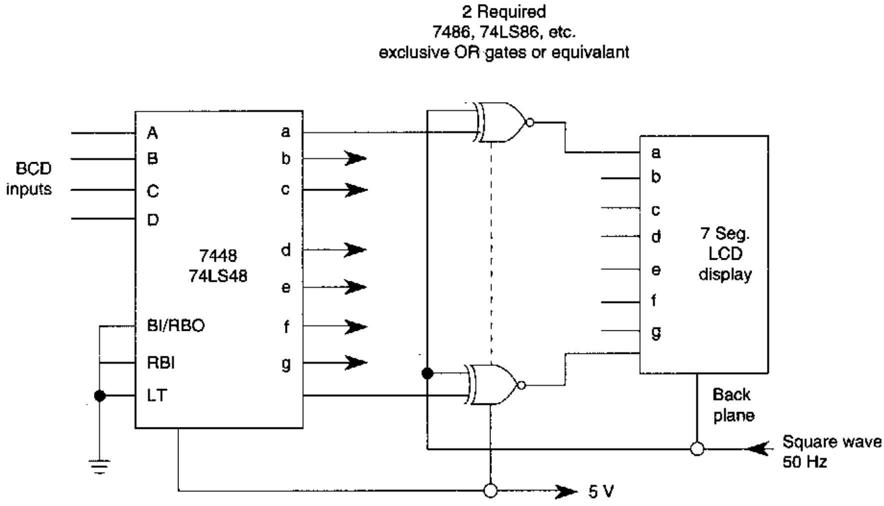
ELECTRONIC DESIGN

FIG. 26-3

Large LCD devices of 1" or more exhibit a large driving capacitance to the driver circuits. To solve this problem, the drive circuit shown (see the figure) introduces a buffer amplifier for each of the three common lines. Each amplifier can be programmed independently for a quiescent current of 10, 100, or 1000 μA. In this application, the bias network applies a voltage that sets the three quiescent currents to 100 μA.

The display driver and triple op amp operate between 5 V and ground, and the COM signals range from 5 V to ≈ 1 V. To ensure that these signals remain within the amplifiers' common-mode range, the signals are attenuated by one-half and the buffers operate at a gain of two. The circuit drives eight 1-inch displays, and is suitable for ambient temperature variations of 15°F or less. At the highest expected temperature, R1 should be adjusted so that no "off" segments are visible.

7-SEGMENT LCD DRIVER

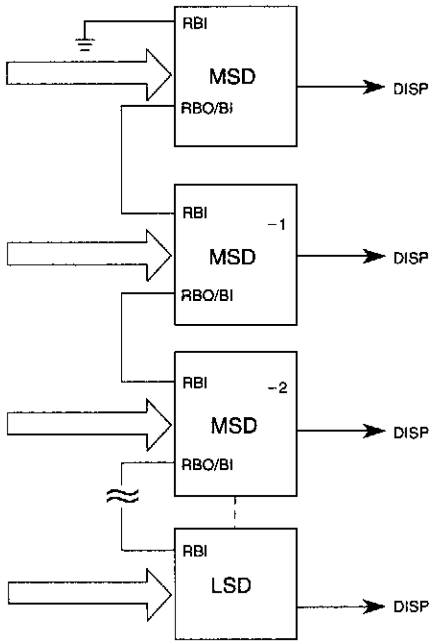


WILLIAM SHEETS

FIG. 26-4

This circuit shows how a 7448 IC is used to drive a 7-segment LCD display. An external 50-Hz square wave supplies necessary phase signals to the back plane of the display.

LED DISPLAY LEADING-ZERO SUPPRESSOR

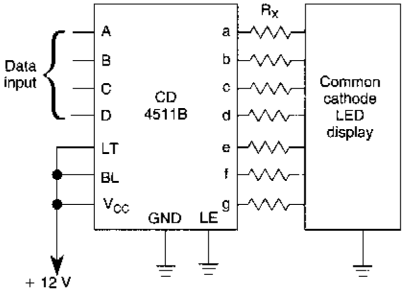


The diagram shows how to connect 7447-type IC devices for leading-zero suppression in an LED display.

WILLIAM SHEETS

FIG. 26-5

7-SEGMENT COMMON-CATHODE LED DISPLAY DRIVER



R_X should limit drive current to segment rated value

$$R_X = \frac{V_{CC} - V_{DISPLAY}}{I_{SEGMENT}}$$

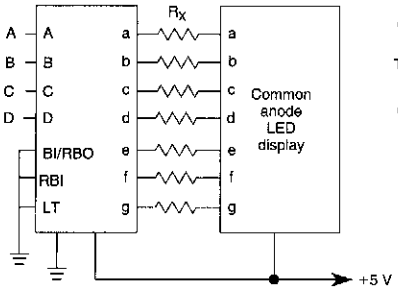
Typical values V_{CC} = 12 V, $V_{DISPLAY}$ = 2.5 V, I_{SEG} = 20 mA

$$R_X = \frac{12 - 2.5}{0.02} = 475\ \Omega \text{ (use 470 } \Omega\text{)}$$

WILLIAM SHEETS ***FIG. 26-6***

A CD4511B CMOS LED display driver can be used to drive a common cathode LED display. Current limiting resistors limit the segment current to the rated value at maximum supply voltage. A sample calculation is shown.

7-SEGMENT (LED) DISPLAY DRIVER



$$R_X = \frac{V_{OUT} - V_{SEGMENT}}{I_{SEGMENT}}$$

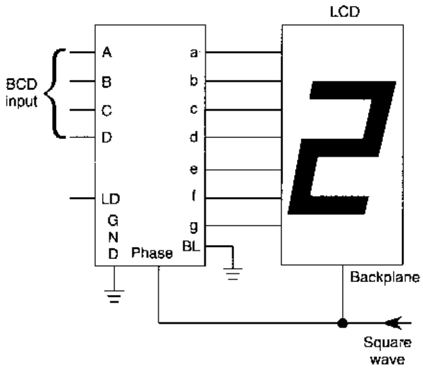
Typical : V_{OUT} = 4.3 V $V_{SEGMENT}$ = 2.5 V I_{SEG} = 20 mA

$$R_X = \frac{4.3 - 2.5}{0.02} = 180\ \Omega$$

An IC1 like a 7447 drives a 7-segment common anode LED display. Current limiting resistor R should limit the segment current to the rated value at maximum supply voltage. A sample calculation is shown.

WILLIAM SHEETS ***FIG. 26-7***

4543B 7-SEGMENT LCD DRIVER



The circuit shows a frequently-used method of driving an LCD display. A square-wave drive is necessary for this application.

WILLIAM SHEETS ***FIG. 26-8***

GAS DISCHARGE TUBE OR DISPLAY DRIVER

WILLIAM SHEETS

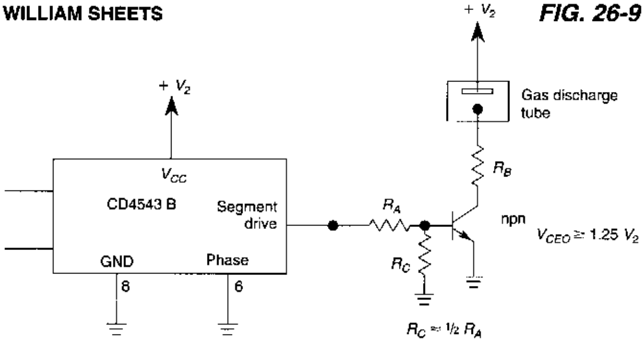
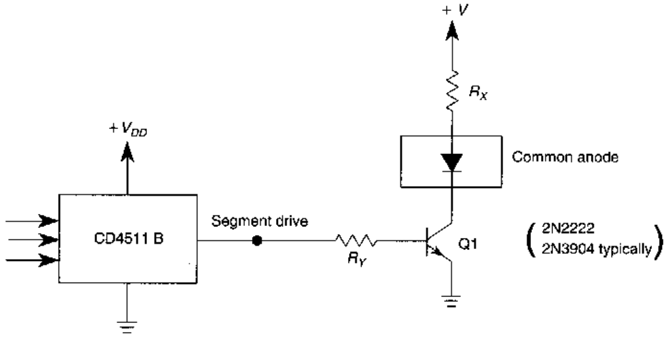


FIG. 26-9

To drive the display, R_A should provide a drive of about 1 mA to the gas discharge tube. R_B is a current-limiting resistor.

4511B COMMON-ANODE DISPLAY DRIVER

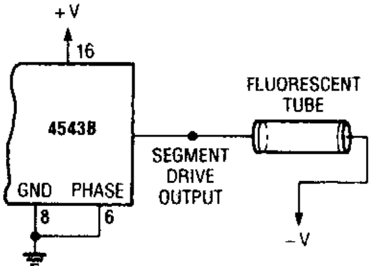


The use of a switching transistor (like a 2N2222 or 2N3904) allows use of the CD4511B with a common-anode display. R_y should be chosen to provide about 1 mA to drive Q1 and R_x should provide enough current to drive the display. For this circuit, the transistor gain (H_{FE}) should be at least the ratio of the segment drive current to the current through R_y.

WILLIAM SHEETS

FIG. 26-10

FLUORESCENT TUBE DISPLAY DRIVER

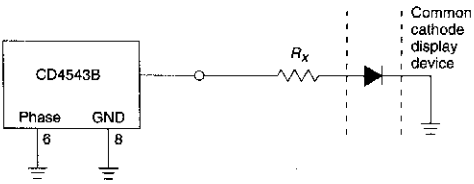


A fluorescent tube or display can be driven with a 4543B IC, as shown.

RADIO-ELECTRONICS

FIG. 26-11

4543B COMMON-CATHODE LED DRIVER



This circuit shows a way of driving a common-cathode display segment or an LED with a CD4543B.

WILLIAM SHEETS

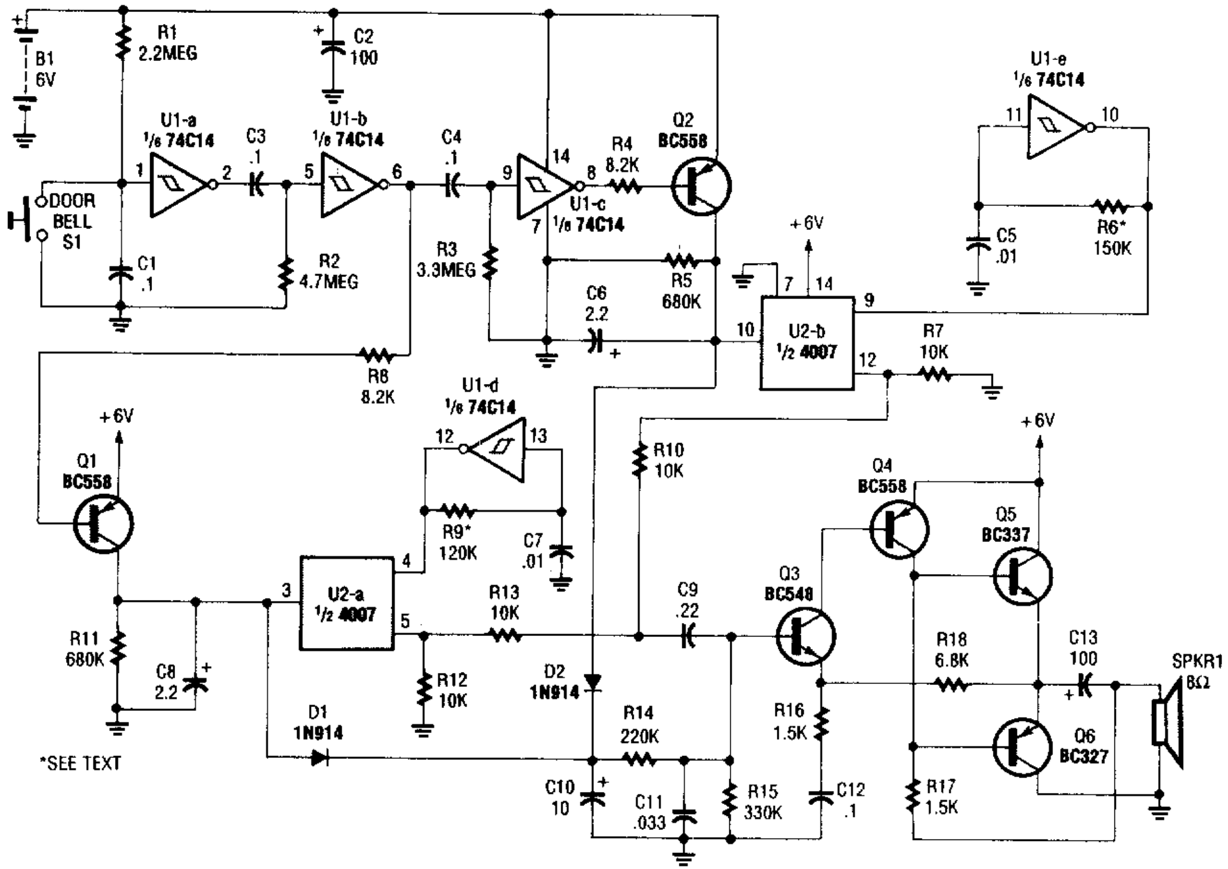
FIG. 26-12

27

Doorbell Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC DOORBELL

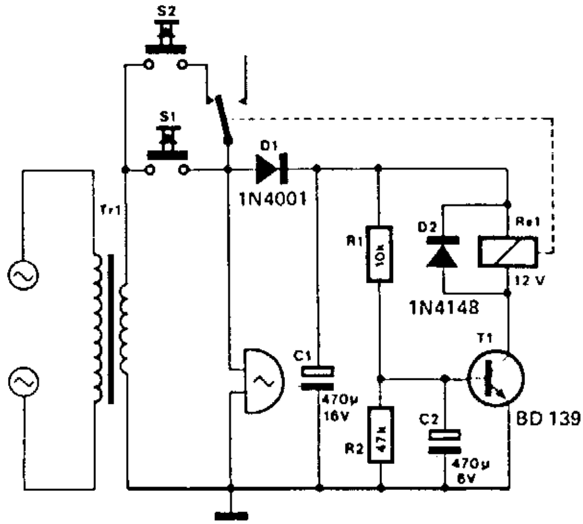


POPULAR ELECTRONICS

FIG. 27-1

When the doorbell switch is pressed, the two monostable stages are activated in sequence, applying bias to a pair of voltage-controlled resistor stages. These then modulate the outputs from a pair of tone generators. The resulting signals are fed to an audio amplifier, then to the speaker.

TWIN BELL CIRCUIT



Tr1 = bell transformer

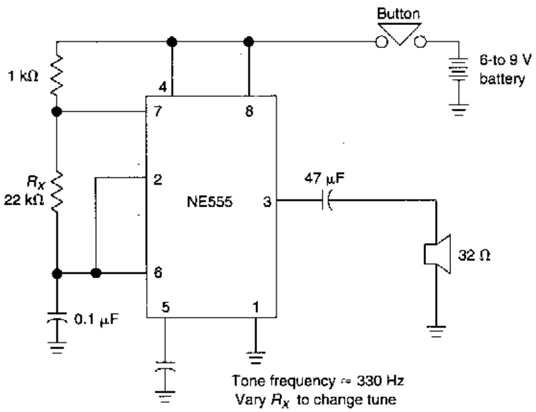
303 CIRCUITS **FIG. 27-2**

It is often desirable for a single doorbell to be operated by two buttons, for instance, one at the front door and the other at the back door.

The additional button, S2 in series with the break contact of relay Re1, is connected in parallel with the original bell-push, S1. When S2 is pressed, the bell voltage is rectified by D1 and smoothed by C1. After a time, t = $R_1R_2C_2$, the direct voltage across C2 has risen to a level here T1 switches on. Relay Re1 is then energized and its contact breaks the circuit of S2 so that the bell stops ringing. After a short time, C1 and C2 are discharged, the relay returns to its quiescent state and the bell rings again.

In this way, S1 will cause the bell to ring continuously, while S2 makes it ring in short bursts, so that it is immediately clear which button is pushed.

ELECTRONIC DOOR BUZZER



This simple electronic door buzzer draws no quiescent current. When S1 is pressed the speaker produces a tone. The NE555 (U1) generates signal.

POPULAR ELECTRONICS **FIG. 27-3**

28

Fax Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Fax Mate

FAX MATE

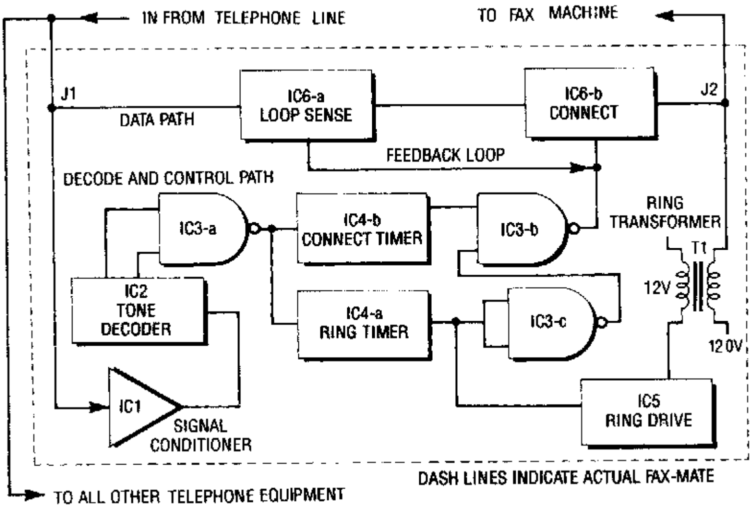


FIG. 1—BLOCK DIAGRAM for the Fax-Mate. The upper path is for data, and the lower one is the decode and control path.

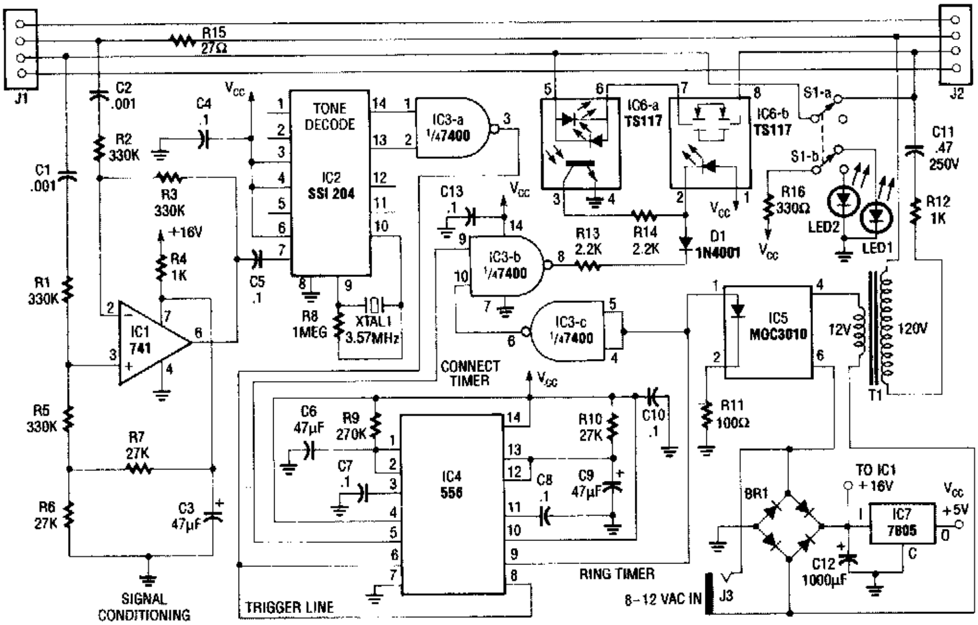


FIG. 2—SCHEMATIC for the Fax-Mate. Notice how it closely resembles the block diagram.

FAX MATE(Cont.)

The fax mate separates the fax machine from the phone line, rings the fax machine on command, connects equipment to incoming lines, and senses the end of the message. When a touch tone pound signal (#) is detected, it actuates a ring greater and driver for the fax machine (the # signal is not used in ordinary dialing). The connect signal is inhibited for this time (ring cycle). 1C46 runs for 15 s and drives part of the connect IC. Then the fax or modem has fired up and is sending out a handshake tone. IC6 connects the equipment for initial hookup and keeps the connect section powered. When the fax machine hangs up, the loop current detector turns off, and resets the system.

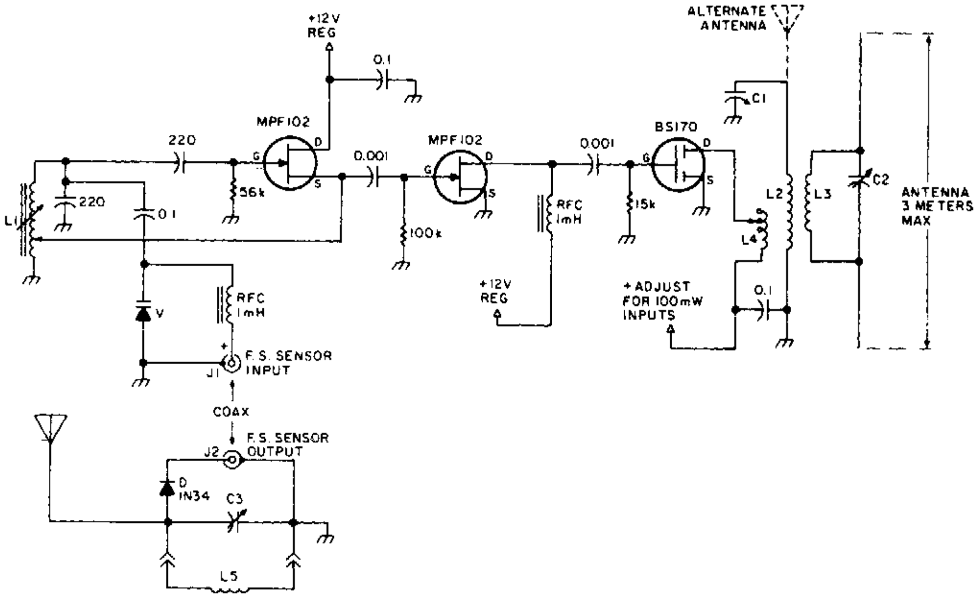
29

Field-Strength Meter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Remote Field Strength Meter
Amplified Field Strength Meter
Simple Amplified Field Strength Meter
Simple Field Strength Meter I
Simple Field Strength Meter II

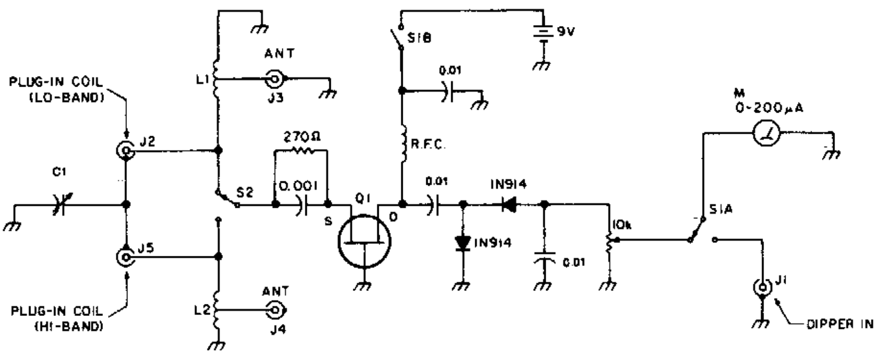
REMOTE FIELD STRENGTH METER



73 AMATEUR RADIO TODAY ***FIG. 29-1***

This field strength meter consists of a tuned crystal detector producing a dc output voltage from a transmitted signal. The dc voltage is used to shift the frequency of a transmitter of 100-mW power operating at 1650 kHz. The frequency shift is proportional to the received field strength. This unit has a range of several hundred feet and is operated under FCC part 15 rules (100-mW max power into a 2-m-long antenna between 510 and 1705 kHz).

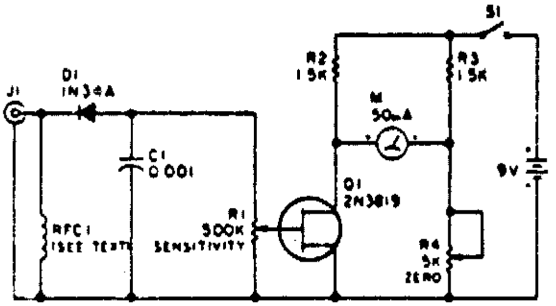
AMPLIFIED FIELD STRENGTH METER



73 AMATEUR RADIO TODAY ***FIG. 29-2***

FET Q1 acts as an RF amplifier to boost sensitivity of the usual diode detector field strength meter.

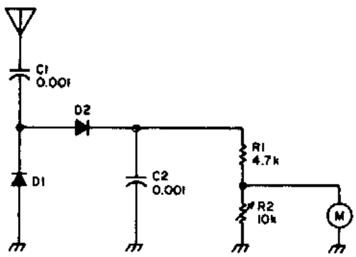
SIMPLE AMPLIFIED FIELD STRENGTH METER



This circuit uses a FET as a dc amplifier in a bridge circuit. R4 is set for meter null with J1 short circuited. Any surplus 50-mA meter can serve in this circuit. RFC1 is any suitable RF choke for the band in use. A 2.5-mH RF choke will do for broadband operation. R1 is a sensitivity control. The antenna can be any small whip antenna (2 ft or less).

73 AMATEUR RADIO TODAY *FIG. 29-3*

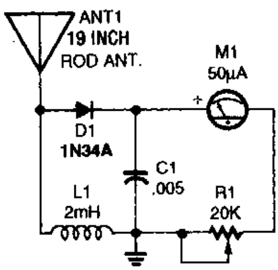
SIMPLE FIELD STRENGTH METER I



Useful for checking transmitters and antennas, this circuit uses a voltage-doubling detector D1 and D2 (HP 5082-2800 hot carrier types). D1 and D2 can also be type IN34 or IN82. M is a 100-mA meter movement.

73 AMATEUR RADIO TODAY *FIG. 29-4*

SIMPLE FIELD STRENGTH METER II



This simple field-strength meter provides a cheap way to monitor an amateur radio or CB transmitter (or even an antenna system) for maximum output.

POPULAR ELECTRONICS *FIG. 29-5*

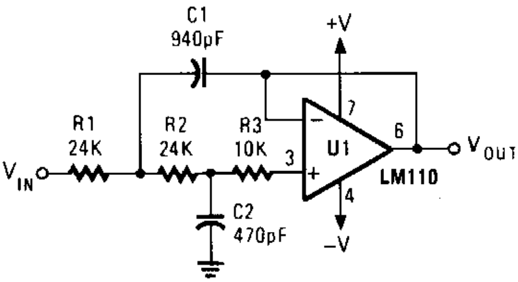
30

Filter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Active Low-Pass Filter
High Q Notch Filter
Universal Stale Variable Filter
Adjustable Q Notch Filter
Fourth Order High-Pass Butterworth Filter
Tunable Notch Filter
High Q Bandpass Filter
Simulated Inductor
Bandpass Filter
Fourth Order Low-Pass Butterworth Filter
Active High-Pass Filter
400-Hz Low-Pass Butterworth Filter
Bandpass Filter
Active Low-Pass RC Filter
Passive L Filter Configurations
Passive Pi Filter Configurations
Four-Output Filter
Variable Q Filter for 400 Hz
Twin T Notch Filter for 1 kHz
Variable Bandpass Audio Filter
Active Fourth-Order Low-Pass Filter
Audio Notch Filter for Shortwave Receivers
Active Second-Order Bandpass Filter
Variable-Frequency Audio BP Filter
Variable Low-Pass Filter
Variable High-Pass Filter
1-mV Offset, Clock-Tunable, Monolithic 5-Pole Low-Pass Filter
Unity-Gain Second-Order High-Pass Filter
Active Unity-Gain Second-Order Low-Pass Filter
Active Fourth-Order High-Pass Filter for 50 Hz
Simple High-Pass (HP) Active Filter for 1 kHz
Equal Second-Order HP Filter
Second-Order Low-Pass Filter for 10 kHz
Simple Low-Pass (LP) Active Filter for 1 kHz
Current-Driven Sallen Key Filter
455-kHz Narrow-Band IF Filter
Audio-Range Filter
BI-Quad RC Bandpass Filter
Passive T Filter Configurations
Full-Wave Rectifier/Averaging Filter
1-kHz Tone Filter

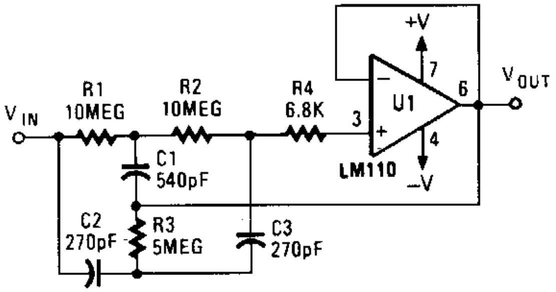
ACTIVE LOW-PASS FILTER



POPULAR ELECTRONICS

FIG. 30-1

HIGH Q NOTCH FILTER

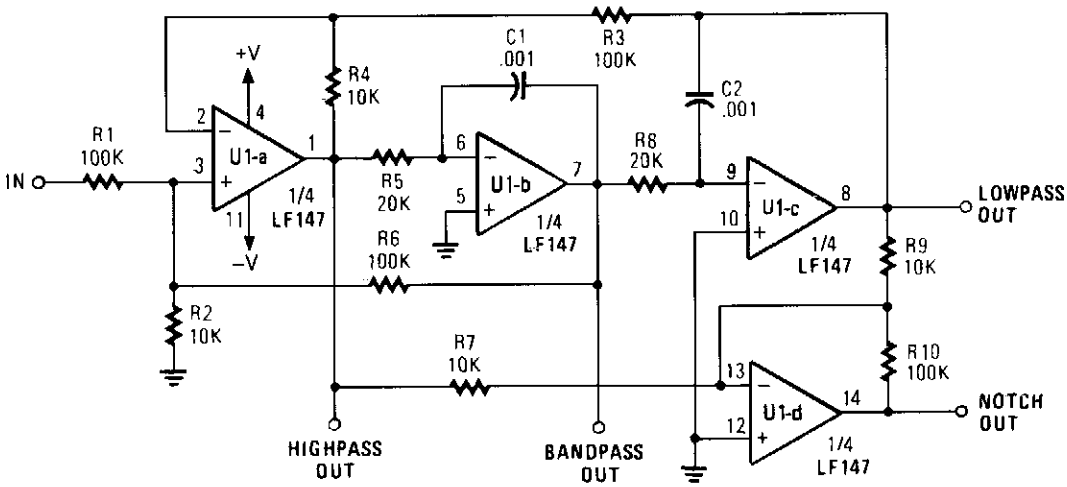


$$R_1 = R_2 = 2R_3 \qquad F_0 = \frac{1}{2\pi R_1, C_2}$$
$$C_2 = C_3 = \frac{C_1}{2}$$

POPULAR ELECTRONICS

FIG. 30-2

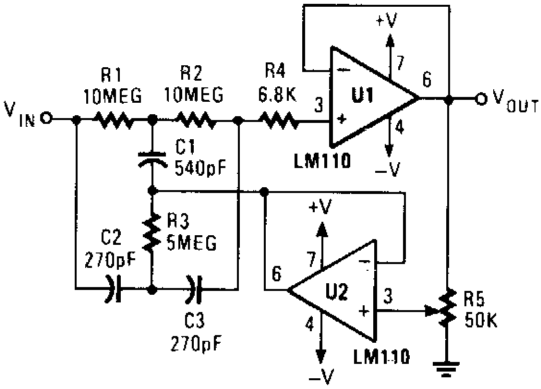
UNIVERSAL STALE VARIABLE FILTER



POPULAR ELECTRONICS

FIG. 30-3

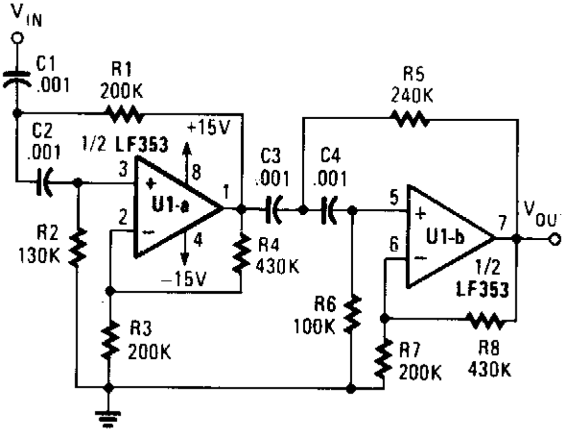
ADJUSTABLE Q NOTCH FILTER



POPULAR ELECTRONICS

FIG. 30-4

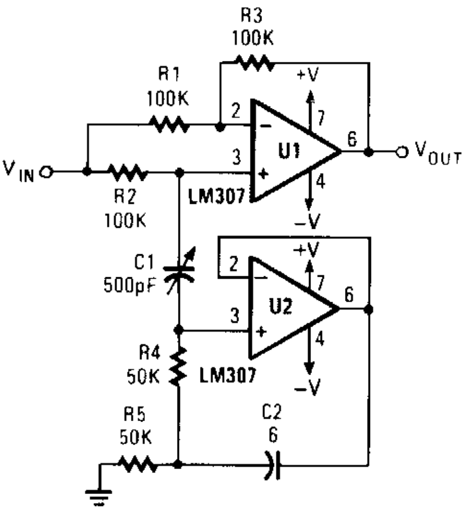
FOURTH ORDER HIGH-PASS BUTTERWORTH FILTER



POPULAR ELECTRONICS

FIG. 30-5

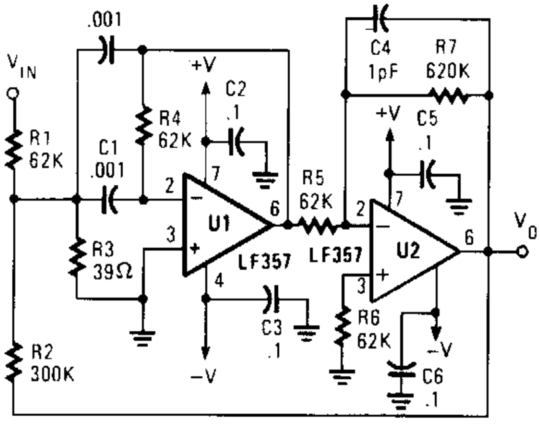
TUNABLE NOTCH FILTER



POPULAR ELECTRONICS

FIG. 30-6

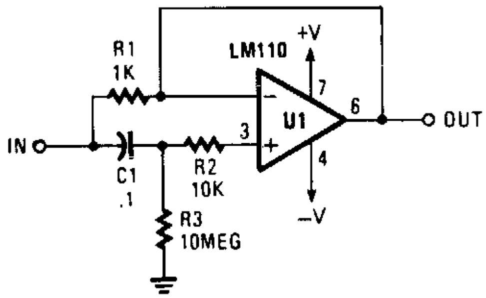
HIGH Q BANDPASS FILTER



POPULAR ELECTRONICS

FIG. 30-7

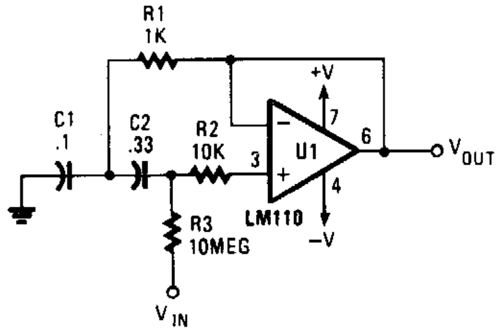
SIMULATED INDUCTOR



POPULAR ELECTRONICS

FIG. 30-8

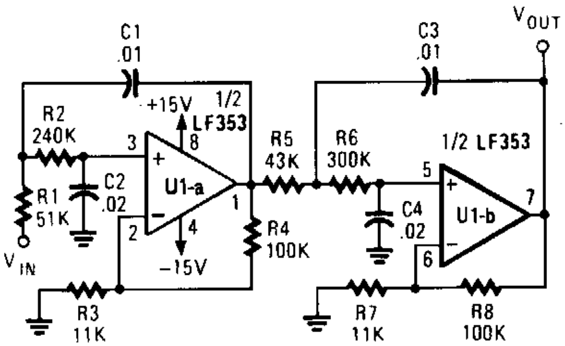
BANDPASS FILTER



POPULAR ELECTRONICS

FIG. 30-9

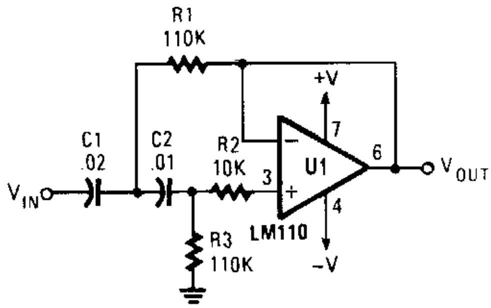
FOURTH ORDER LOW-PASS BUTTERWORTH FILTER



POPULAR ELECTRONICS

FIG. 30-10

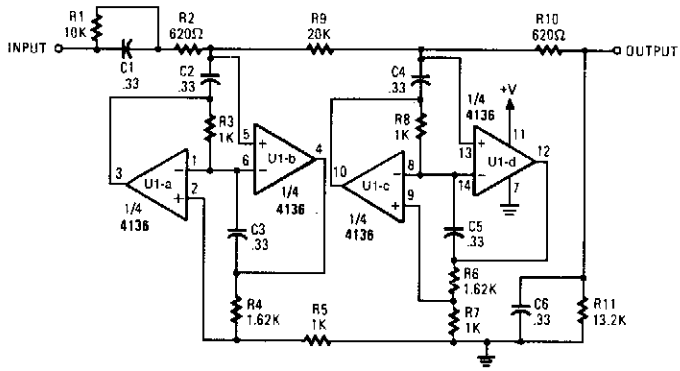
ACTIVE HIGH-PASS FILTER



POPULAR ELECTRONICS

FIG. 30-11

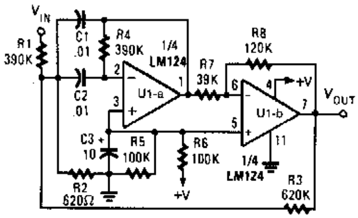
400-Hz LOW-PASS BUTTERWORTH FILTER



POPULAR ELECTRONICS **FIG. 30-12**

Designed for a 400-Hz cutoff frequency, the cutoff can be scaled by varying the element values proportionally to frequency

BANDPASS FILTER



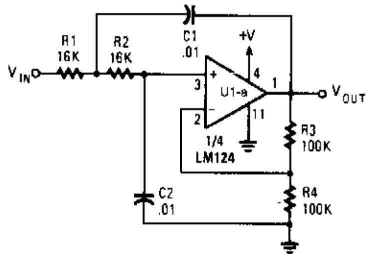
POPULAR ELECTRONICS **FIG. 30-13**

Appropriate center frequency of this circuit is:

$$\frac{1}{R_4 C_2}$$

$$C_1 = C2,\ R_1 = R_4$$

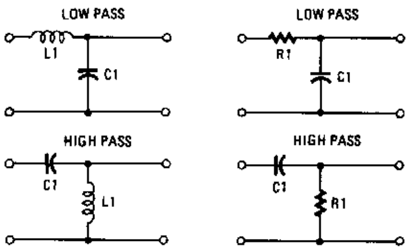
ACTIVE LOW-PASS RC FILTER



POPULAR ELECTRONICS **FIG. 30-14**

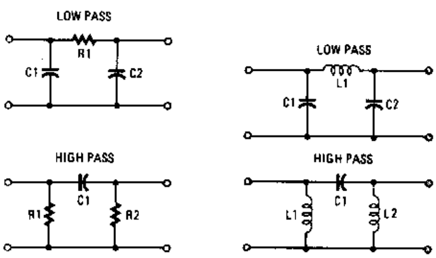
The circuit shown has a cutoff frequency at about 1 kHz. R1, R2, C1, and C2 can be scaled to change this to any other desired frequency.

PASSIVE L FILTER CONFIGURATIONS



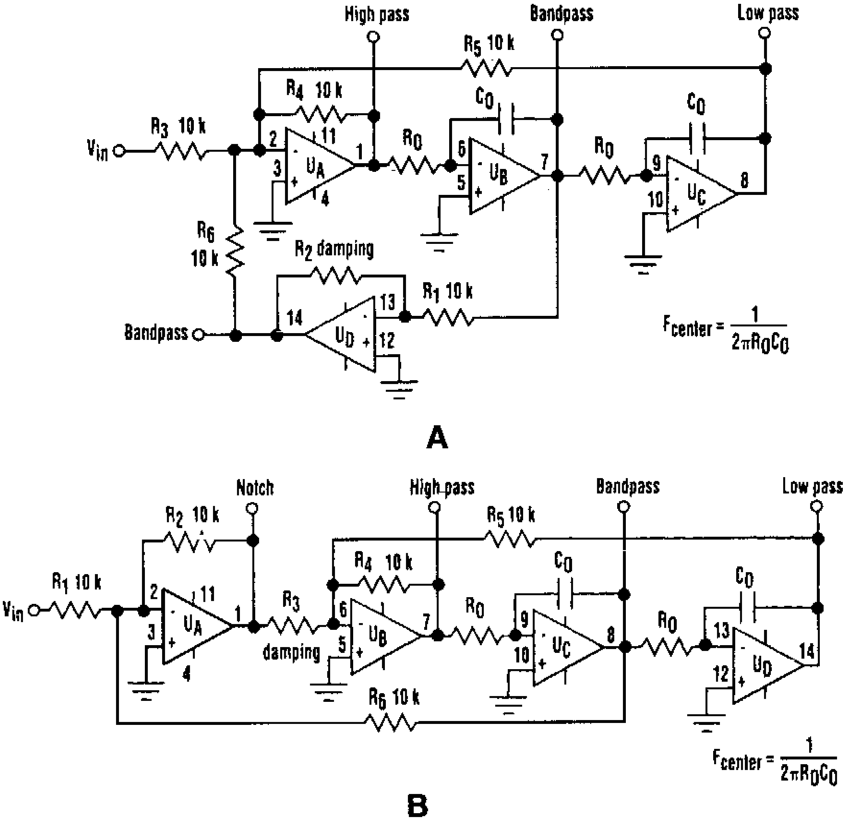
POPULAR ELECTRONICS **FIG. 30-15**

PASSIVE PI FILTER CONFIGURATIONS



POPULAR ELECTRONICS **FIG. 30-16**

FOUR-OUTPUT FILTER



ELECTRONIC DESIGN

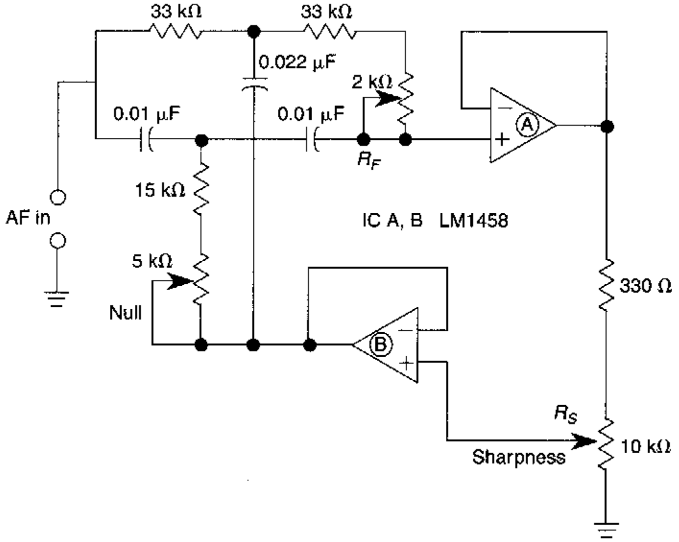
FIG. 30-17

The classic "state-variable" (two-integrator) filter (see Fig. A) is famous for its insensitivity to device parameter tolerances, as well as its ability to provide three simultaneous separate outputs: high pass, bandpass, and low pass. These advantages often offset the fact that a quad operational amplifier is needed to implement the circuit.

A modification of the classic scheme that applies the input voltage via amplifier U_D, rather than U_A provides a bandpass output with a fixed peak gain that doesn't depend on the Q of the filter. It was found by using that configuration, a fourth notch-filter output can be obtained if $R_1 = R_6$ (see Fig. B).

If $R_1 = R_6 = R_2$, the gains of both the notch and bandpass outputs are unity, regardless of the Q factor, as determined by R3, R1, R2, R4, R5, and R6. The resonant (or cutoff) frequency is given by ω, – $1/R_O \times C_O$. Depending on the capacitor values and frequency ω, resistance R_O might also share the same monolithic network for maximum space economy. As with the classic configuration, resonant frequency ω can be electrically controlled by switching resistors R_O, or by using analog multipliers in series with the integrators.

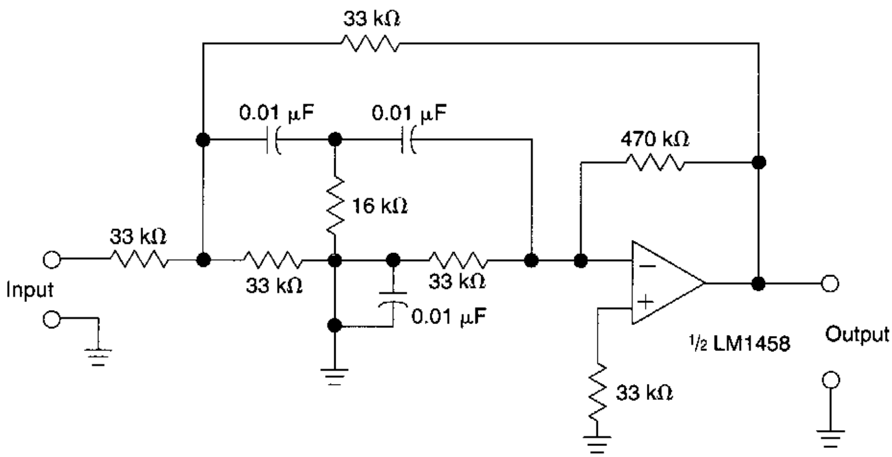
VARIABLE *Q* FILTER FOR 400 Hz



WILLIAM SHEETS ***FIG. 30-18***

A bootstrapped twin T notch filter in this circuit can yield an effective Q of up to 10. R_S adjusts the feedback, hence the Q. Values of C_1 and C_2 can be changed to alter the frequency. R_F is a fine-tune null control.

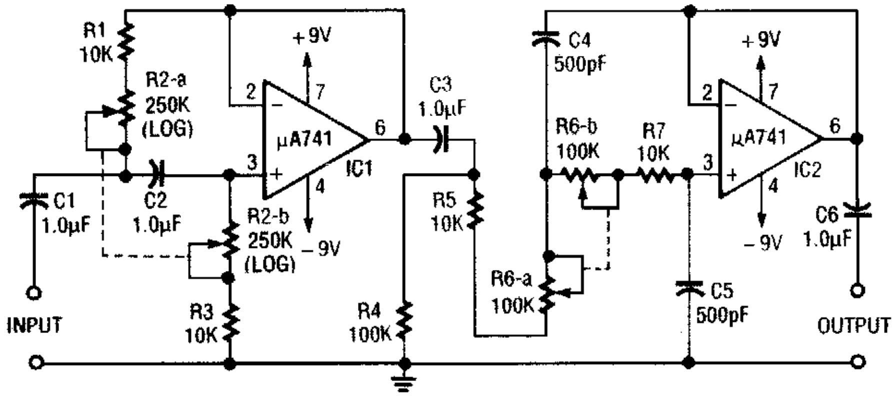
TWIN T NOTCH FILTER FOR 1 kHz



WILLIAM SHEETS ***FIG. 30-19***

The circuit shown uses a twin T notch filter and an amplifier. Used to remove unwanted frequency.

VARIABLE BANDPASS AUDIO FILTER

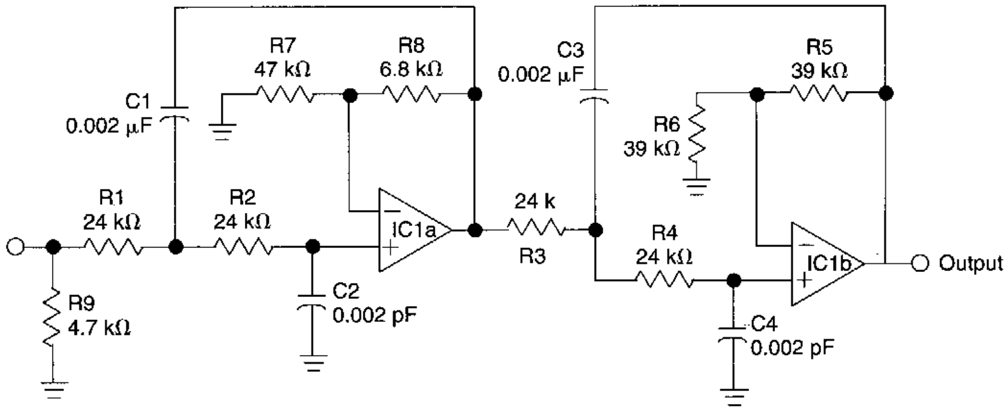


ELECTRONICS NOW

FIG. 30-20

This circuit is a variable audio bandpass filter that has a low cutoff variable from about 25 Hz to 700 Hz and a high cutoff variable from 2.5 kHz to over 20 kHz. Rolloff is 12 dB/octave on both high and low ends. R2-a-b and R6-a-b are ganged potentiometers for setting lower and upper cutoff frequencies, respectively.

ACTIVE FOURTH-ORDER LOW-PASS FILTER



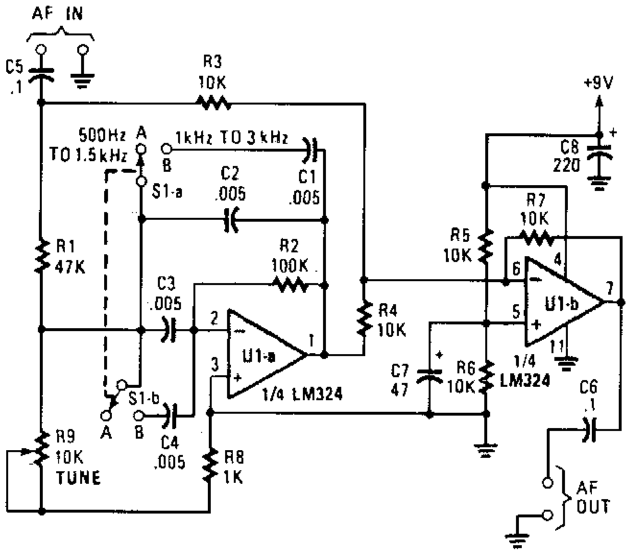
IC1 a, b op amp = LM1458

WILLIAM SHEETS

FIG. 30-21

This circuit is a fourth-order low-pass filter with values for kHz. The values of R_1, R_2, C_1 and C_2, and R_3, R_4, C_3 and C_4 can be scaled for operation at other frequencies. Roll-off is 24 dB/octave.

AUDIO NOTCH FILTER FOR SHORTWAVE RECEIVERS

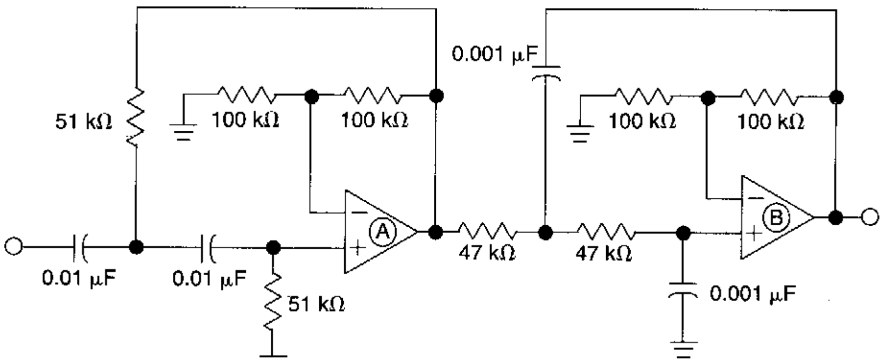


POPULAR ELECTRONICS

FIG. 30-22

The notch filter can be added to just about any receiver to attenuate a single frequency by more than 30 dB. This filter should be handy for reducing heterodynes and whistles.

ACTIVE SECOND-ORDER BANDPASS FILTER FOR SPEECH RANGE



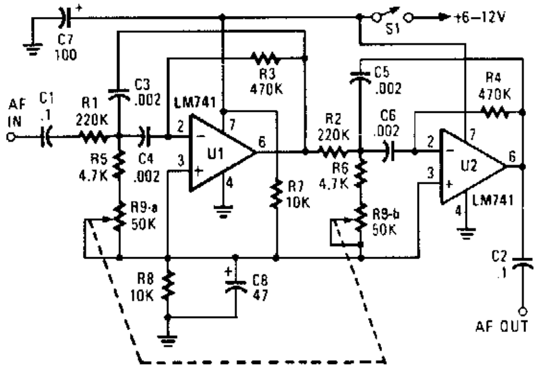
Op amp LM1458 dual, etc.

WILLIAM SHEETS

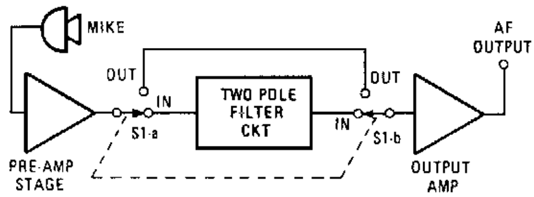
FIG. 30-23

This filter circuit which uses LM1458 or similar op amp has a response of 300 Hz to 3.4 kHz with 12 dB/octave roll-off outside the pass band. Section A is the high-pass one, followed by low-pass section B. Values of either section can be scaled to alter the pass band.

VARIABLE-FREQUENCY AUDIO BP FILTER



A

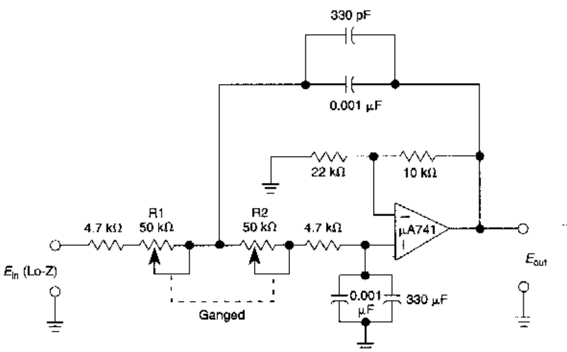


The filter can be wired into an existing amplifier by inserting the filter circuit between the amp's preamp and output stages as shown here.

POPULAR ELECTRONICS ***FIG. 30-24***

This variable-frequency, audio bandpass filter is built around two 741 op amps that are connected in cascade. Two 741 op amps are configured as identical RC active filters and are connected in cascade for better selectivity. The filter's tuning range is from 500 Hz to 1500 Hz. The overall voltage gain is slightly greater than 1 and the filter's is about 5. The circuit can handle input signals of 4 V peak-to-peak without being overdriven. The circuit's input impedance is over 200 kΩ and its output impedance is less than 1 kΩ .

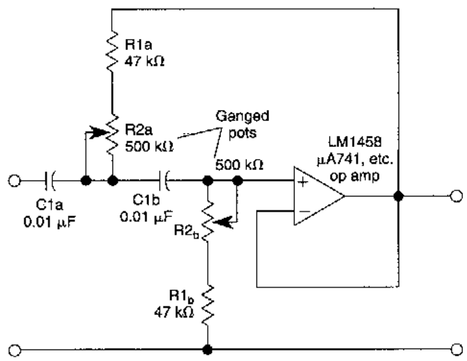
VARIABLE LOW-PASS FILTER



WILLIAM SHEETS ***FIG. 30-25***

This second-order low-pass filter uses a 741 op amp and is tuneable from 2.5 kHz to 25 kHz. This circuit is useful in audio and tone control applications. R1 and 2 are ganged potentiometers.

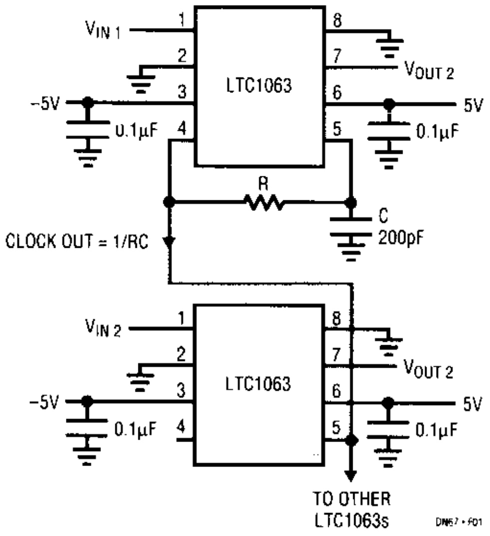
VARIABLE HIGH-PASS FILTER



WILLIAM SHEETS ***FIG. 30-26***

This second order filter which should prove useful in audio applications uses an LM1458 or other similar of op amp. It is tuneable from 30 to 300 Hz cutoff. R2a, b are ganged log-taper potentiometers.

1-mV OFFSET, CLOCK-TUNABLE, MONOLITHIC 5-POLE LOW-PASS FILTER



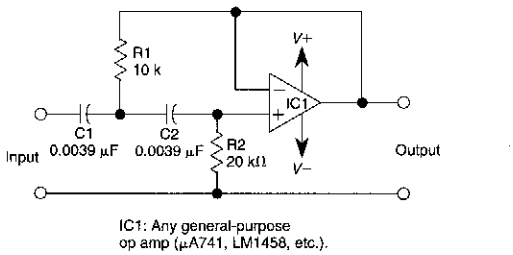
The LTC1063 is the first monolithic low-pass filter that simultaneously offers outstanding dc and ac performance. It features internal or external clock tunability, cutoff frequencies up to 50 kHz, 1-mV typical output dc offset, and a dynamic range in excess of 12 bits for over a decade of input voltage.

The LTC1063 approximates a 5-pole Butterworth low-pass filter. The unique internal architecture of the filter allows outstanding amplitude matching from device to device. Typical matching ranges from 0.01 dB at 25% of the filter passband to 0.05 dB at 50% of the filter passband.

An internal or external clock programs the filter's cutoff frequency. The clock-to-cutoff frequency ratio is 100:1. In the absence of an external clock, the LTC1063's internal precision oscillator can be used. An external resistor and capacitor set the device's internal clock frequency.

LINEAR TECHNOLOGY CORP. ***FIG. 30-27***

UNITY-GAIN SECOND-ORDER HIGH-PASS FILTER



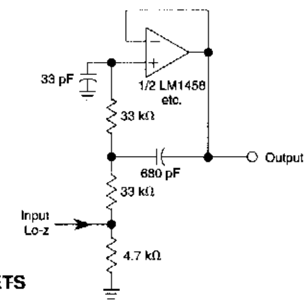
WILLIAM SHEETS ***FIG. 30-28***

This filter circuit has a cutoff frequency of 2900 Hz with the values shown.

$$f_{\text{cutoff}} = \frac{1}{2.83\pi RC}$$

$$R = R_1$$
$$R_2 = 2R_1$$
$$C = C_1 = C_2$$

ACTIVE UNITY-GAIN SECOND-ORDER LOW-PASS FILTER



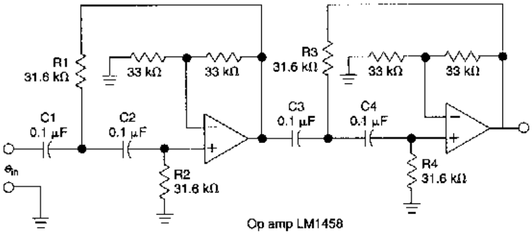
WILLIAM SHEETS ***FIG. 30-29***

This second-order Butterworth filter cuts off near 10 kHz. The values of C_1 and C_2 can be changed to alter the frequency, or else calculated from the formula.

$$f_{\text{cutoff}} = \frac{1}{2.83\pi RC}$$

$$C_1 = 2C_2$$
$$R_2 = R_3 = R$$

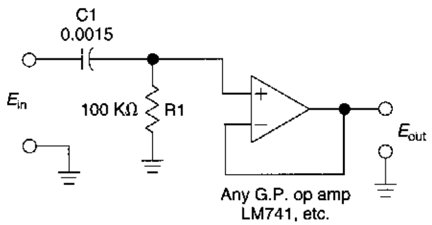
ACTIVE FOURTH-ORDER HIGH-PASS FILTER FOR 50 Hz



This circuit which uses an LM1458 or similar op amp is a fourth-order high-pass filter with a 24 dB/octave roll-off. The values of R_1/R_2, R_3/R_4, C_1/C_2, C_3/C_4 can be scaled to suit other cutoff frequencies.

WILLIAM SHEETS ***FIG. 30-30***

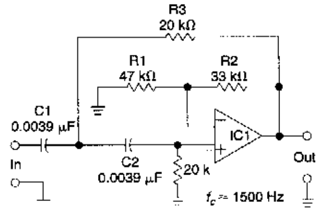
SIMPLE HIGH-PASS (HP) ACTIVE FILTER FOR 1 kHz



WILLIAM SHEETS ***FIG. 30-31***

This simple 1 kHz filter uses a voltage follower and an RC section for a filter element. For other frequencies f_3 dB – 1/6.28 R_1C_1. The response drops 6 dB/octave below f_3 dB.

EQUAL COMPONENTS SECOND-ORDER HP FILTER



IC1: General-purpose op amp (741, etc.).

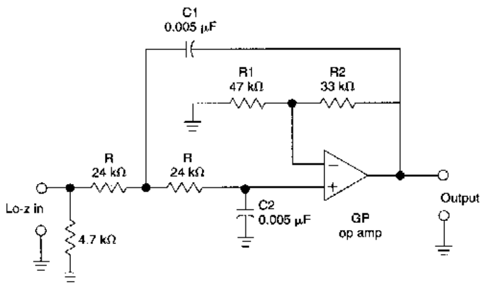
WILLIAM SHEETS ***FIG. 30-32***

This filter circuit uses equal value components and is shown for 1500 Hz. The values can be scaled for other frequencies.

$$f_{\text{cutoff}} = \frac{1}{2.83\pi RC}$$

$$R = R_1$$
$$R_2 = 2R_1$$
$$C = C_1 = C_2$$

SECOND-ORDER LOW-PASS FILTER FOR 10 kHz

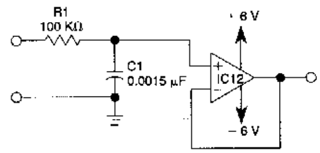


WILLIAM SHEETS ***FIG. 30-33***

This circuit uses equal value capacitors. The cutoff frequency (f_c) is

$$f_c = \frac{1}{2.83\pi RC}$$

SIMPLE LOW-PASS (LP) ACTIVE FILTER FOR 1 kHz

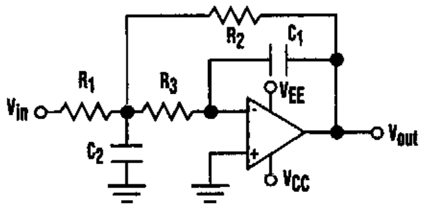


IC1 1/2 LM1458 or general-purpose op amp

WILLIAM SHEETS ***FIG. 30-34***

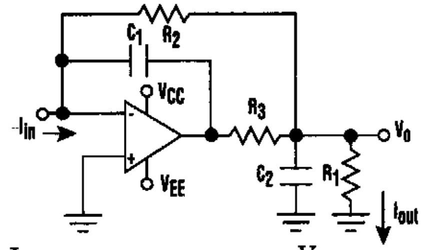
This simple filter uses an RC section for a filter element, with a voltage follower for other frequencies f_3 dB = 1/6.28 R_1C_1. Response drops 6 dB/octave above f_3 dB.

CURRENT-DRIVEN SALLEN KEY FILTER



$$\frac{V_{out}}{V_{in}} = \frac{-1/R_1R_3C_1C_2}{s^2 + s\frac{1}{C_2}\left(\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3}\right) + \frac{1}{R_2R_3C_1C_2}} = H(s)$$

A



$$\frac{I_{out}}{I_{in}} = H(s) \qquad \frac{V_o}{I_{in}} = R_1H(s)$$

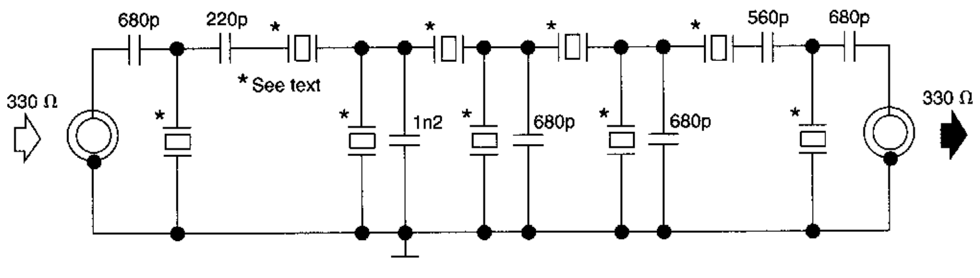
B

ELECTRONIC DESIGN

FIG. 30-35

The low-pass Sallen-Key filter is staple for designers because it contains few components (A). By redesigning the filter, a current to voltage conversion can be avoided when the input signal to be filtered is in current form (B).

455-kHz NARROW-BAND IF FILTER

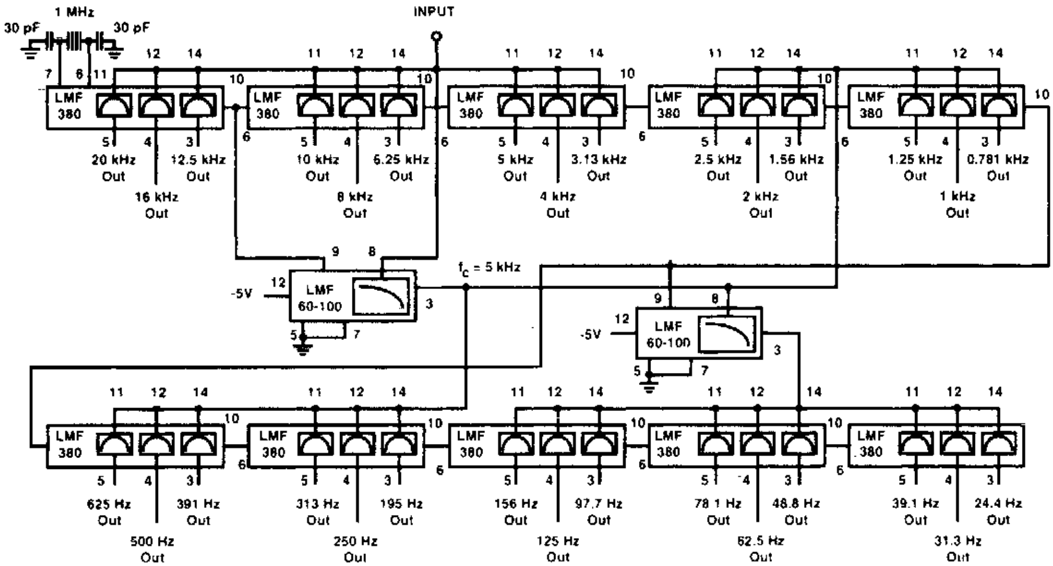


303 CIRCUITS

FIG. 30-36

This filter uses five 455-kHz ceramic resonators. The impedance is 330 Ω, the bandwidth is 800 Hz, and the ultimate rejection ≥60 dB. The ceramic resonators could be replaced by crystals.

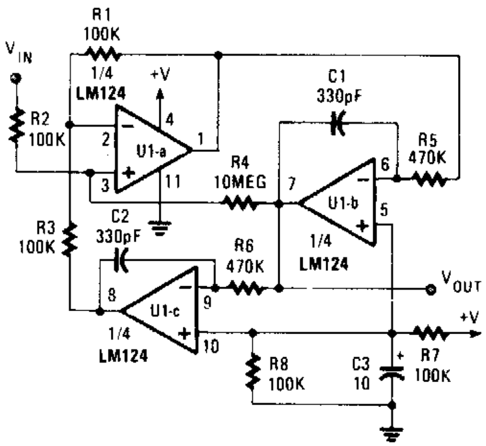
AUDIO-RANGE FILTER



NATIONAL SEMICONDUCTOR *FIG. 30-37*

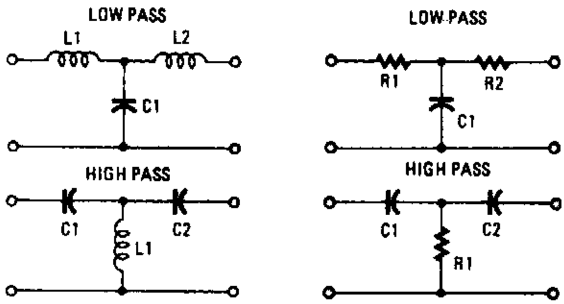
The LMF380 switched audio filter by National Semiconductor is used here to obtain a third-octave filter set that covers the entire audio range.

BI-QUAD RC BANDPASS FILTER



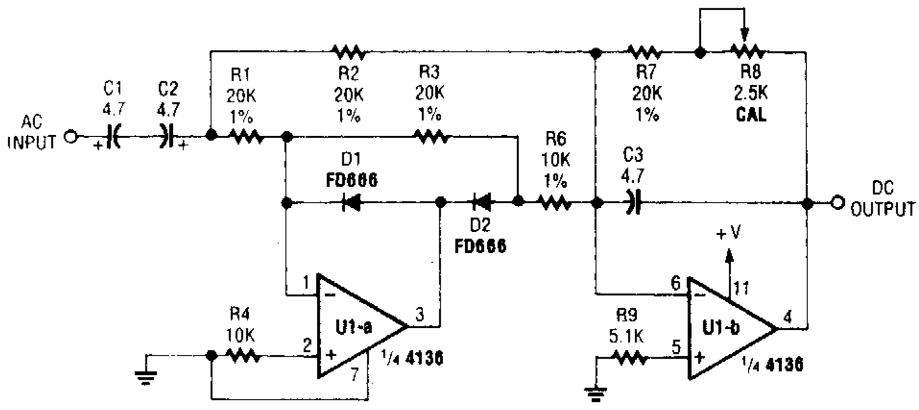
POPULAR ELECTRONICS *FIG. 30-38*

PASSIVE T FILTER CONFIGURATIONS



POPULAR ELECTRONICS *FIG. 30-39*

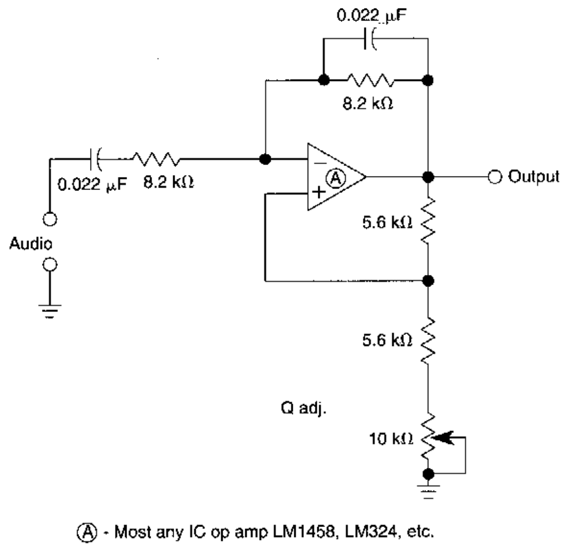
FULL-WAVE RECTIFIER/AVERAGING FILTER



POPULAR ELECTRONICS ***FIG. 30-40***

The input signal is rectified by D1 and D2 op amp U1-a, and fed to output amp U2. R8 is set for correct circuit calibration.

1-kHz TONE FILTER



The Wien-bridge based filter has a variable bandwidth and a center frequency of 900 Hz. The circuit will oscillate if the 10-kΩ pot is set too low.

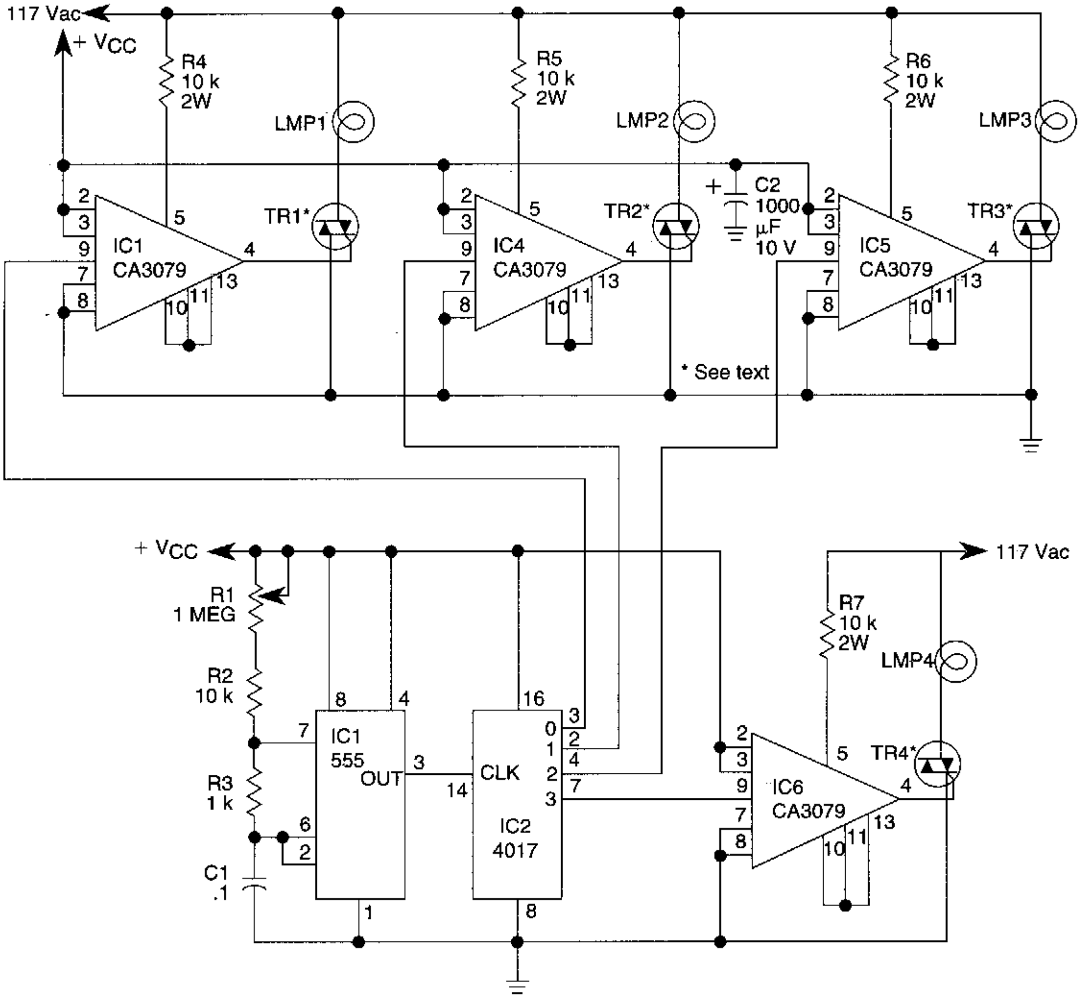
WILLIAM SHEETS ***FIG. 30-41***

31

Flasher Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SEQUENTIAL FLASHER

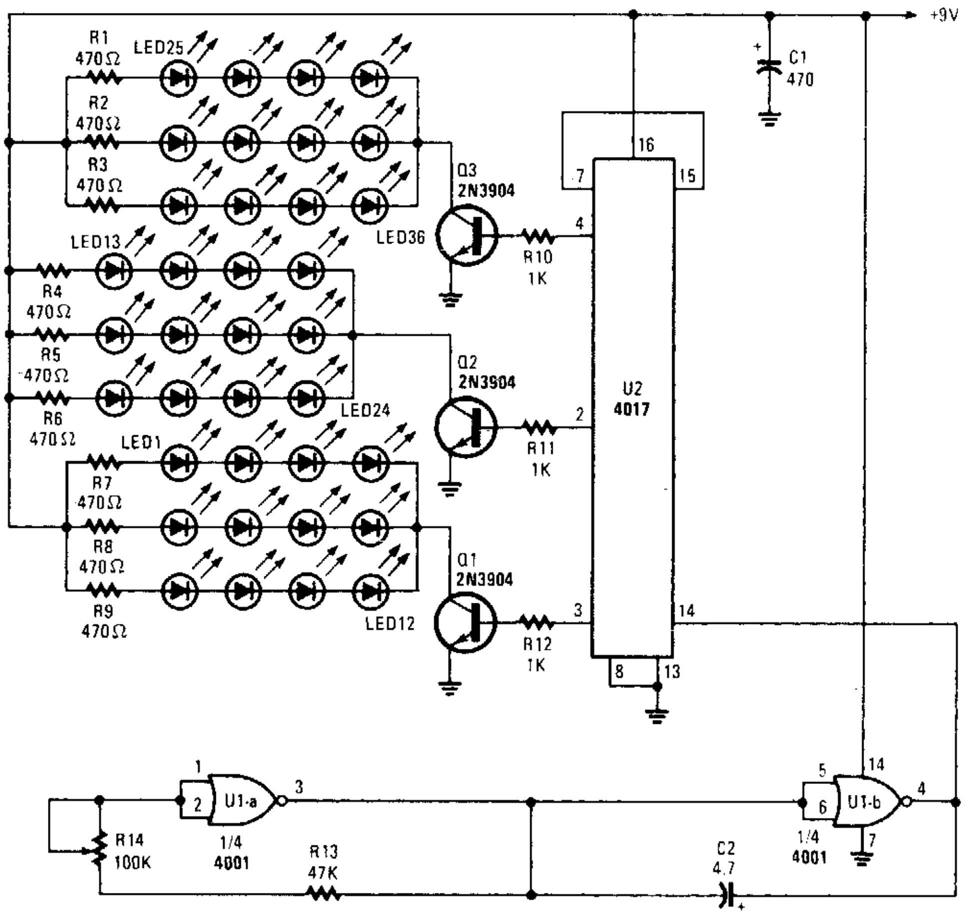


R-E EXPERIMENTERS HANDBOOK *FIG. 31-1*

A 555 timer, IC1, drives a 4017 CMOS decade counter. Each of the 4017's first four outputs drives a CA3079 zero-voltage switch. Pin 9 of the CA3079 is used to inhibit output from pin 4, thereby disabling the string of pulses that the IC normally delivers. Those pulses occur every 8.3 ms, i.e., at a rate of 120 Hz. Each pulse has a width of 120 µs.

Because of the action of the CA3079, the lamps connected to the triacs turn on and off near the zero crossing of the ac waveform. Switching at that point increases lamp life by reducing an inrush of current that would happen if the lamp were turned on near the high point of the ac waveform. In addition, switching at the zero crossing reduces radio frequency interference (RFI) considerably. **Caution:** The CA3079s are driven directly from the 117-Vac power line, so use care.

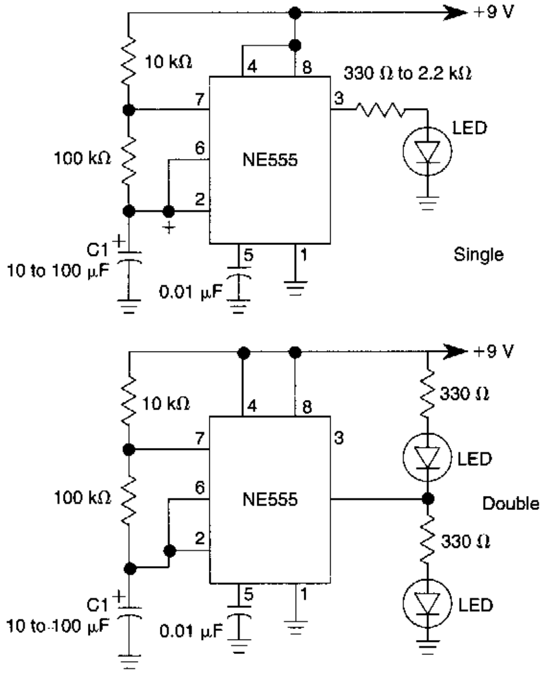
36 LED FLASHER DRIVER



POPULAR ELECTRONICS *FIG. 31-2*

Originally intended as a 3-bell animation circuit for Christmas decorations, the circuit can be used for many other purposes that require a flasher of this kind. By re-connecting U2 (see the data manual), more than three outputs can be be obtained.

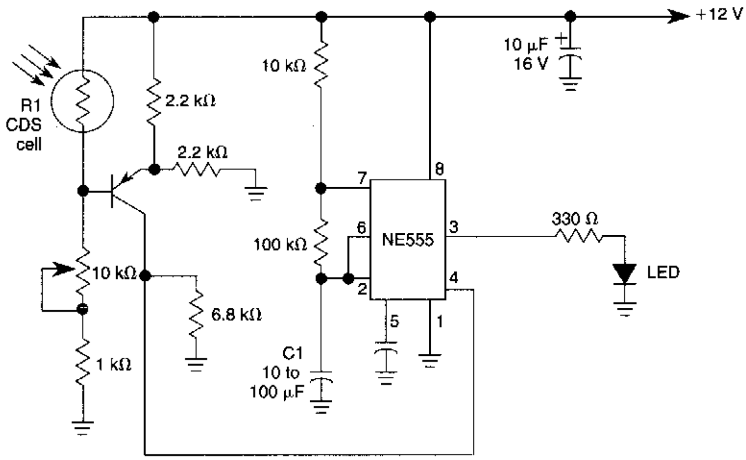
LED FLASHERS



A 555 is used to switch an LED on and off. C1 determines the flash rate. Single ended (one LED) and double-ended (alternating) flashers are shown.

WILLIAM SHEETS *FIG. 31-3*

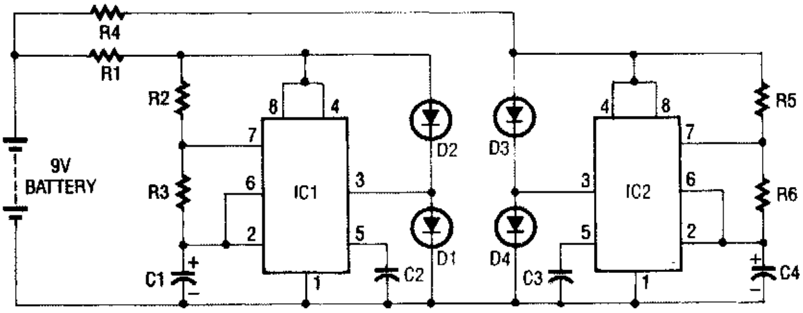
DARK-ACTIVATED LED FLASHER



WILLIAM SHEETS *FIG. 31-4*

This circuit can be used as a small beacon or marker light, and toys or novelty items. R1 is an LDR that has ≥10 kΩ dark-resistance, or a CDS photocell. C1 determines the flash rate.

SUPER LED FLASHER



C1, C4 .... 4.7 μF Electrolytic Capacitor
C2, C3 ..... 330 pF Disc Capacitor
D1 .... Yellow LED
D2, D3 .... Red LED
D4 .... Green LED
IC1, IC2 .... 555 Timer IC
R1, R4 .... 100 ohm Resistor
R2, R5 .... 82 k Resistor
R3, R6 .... 33 k Resistor

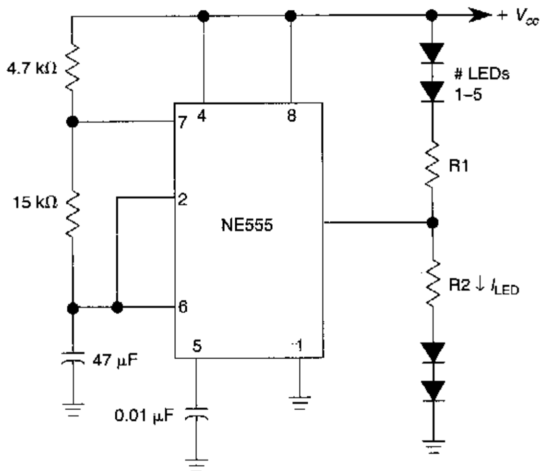
1991 PE HOBBYIST HANDBOOK ***FIG. 31-5***

The super LED flasher is actually two complete LED flasher circuits on one circuit board. The first LED flasher is made up of IC1 and LEDs D1 and D2. IC1 is a 555 timer IC configured as an astable (free-running) multivibrator with its output on pin 3.

The frequency of the 555's oscillation is controlled by R2, R3, and C1. Resistor R1 limits the input voltage to a low enough level to prevent damage to the IC. As the 555 IC oscillates, the output of pin 3 goes high (+) then low (–). When the output is high it supplies current to D1, which lights up. When it is low, pin 3 sinks current and D2 lights up. This happens because LEDs are polarity-sensitive (like all other diodes, they permit current flow in only one direction) and one lead of each LED has been connected to the respective polarity needed to light that LED.

The second LED flasher, made up of IC2 and LEDs D3 and D4, operates in the same way as the first LED flasher.

LED FLASHER FOR 2 TO 10 LEDs



$$R_1, R_2 = \frac{V_{cc} - 2(\#\text{LEDs})}{I_{LED}}$$

Typically $V_{cc} = 12$ V
#LEDs = 2
$I_{LED} = 30$ mA

(# LEDS 1 to 5 per side)

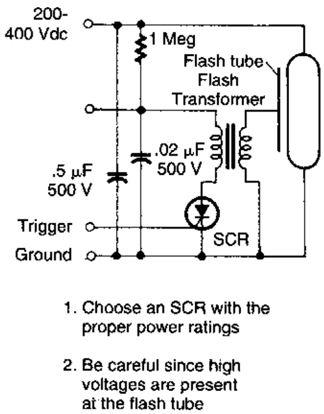
$$R_1, R_2 = \frac{12 - 2(2)}{0.03} \approx 267\ \Omega$$

Use 270 Ω

WILLIAM SHEETS ***FIG. 31-6***

This LED flasher has double-ended output connection. The circuit can be used with 1 to 5 LEDs on each side as indicated.

FLASH SIGNAL ALARM

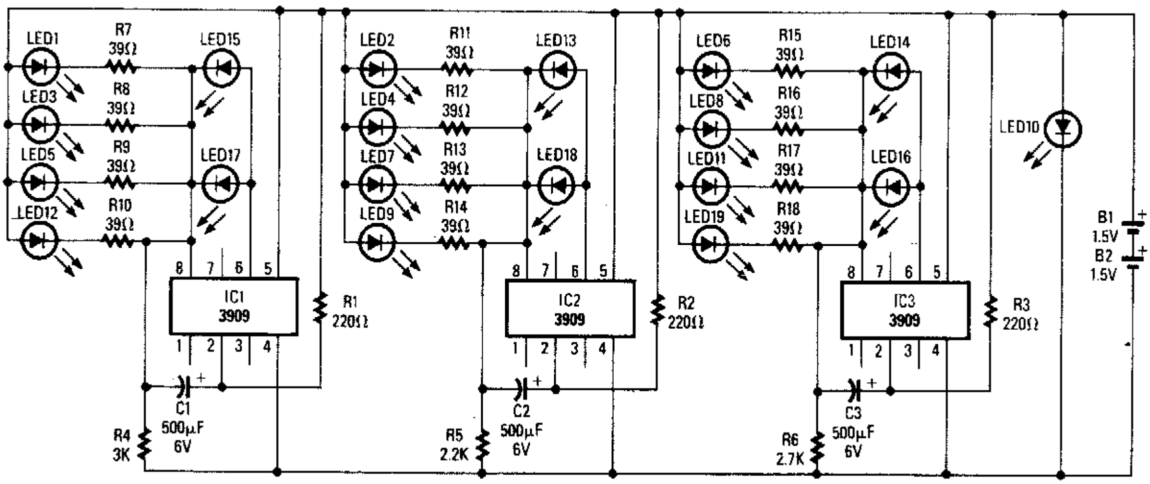


RADIO-ELECTRONICS

FIG 31-7

This circuit is useful if you need a low-energy flashing alarm. The 200 to 400-dc supply should have enough internal resistance to charge the 0.5 μF capacitor between flashes, about 2 or 3 time constants, which means about 500 kΩ to 1 MΩ for a 1-s rate. Use lower values for higher rates.

LED CHRISTMAS TREE LIGHT FLASHER



R-E EXPERIMENTERS HANDBOOK

FIG. 31-8

Three individual flashing circuits that use an LM3909 LED flasher/oscillator IC create the appearance of a pseudo-random firing order. The combination of C_1/R_4, C_2/R_5, and C_3/R_6 control the blink rate, which is between 0.3 and 0.8 s, and the inherent wide tolerance range (–20% to +80%) of standard electrolytic capacitors add to the irregularities of the blink cycles. The continuous current drain is about 10 mA; however, if you decrease the values of R4 through R6 or C1 through C3 in order to increase the blink rate, the current will then increase proportionally.

Note in particular that external current-limiting resistors aren't needed for LED13 through LED18; the resistors are built into the ICs. LED10, which serves as the tree's "star," is a special kind of flashing LED that blinks continuously at a fixed rate.

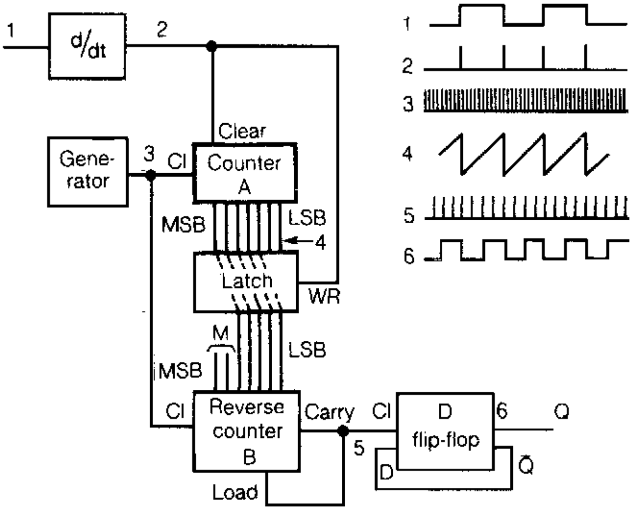
32

Frequency Multiplier Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Frequency Multiplier Without PLL

FREQUENCY MULTIPLIER WITHOUT PLL



ELECTRONIC ENGINEERING

FIG. 32-1

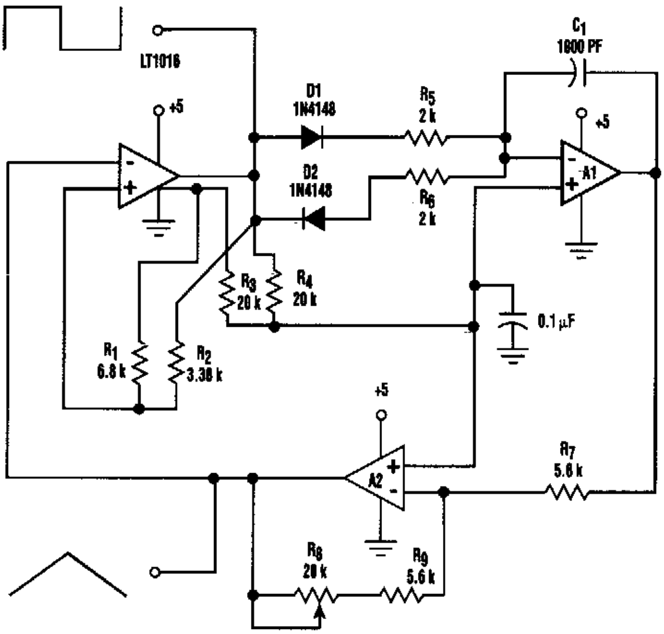
An input rectangular signal is differentiated and short impulses are formed from its edges. These impulses write the content of counter A to a latch that clears the counter after a very short time. Counter A counts impulses of the frequency f_o that are much greater than that of the input signal. The pulses come from an impulse generator. Thus, the number, which is written to the latch, expresses the number of these impulses between the edges of the input signal. The impulses from the same generator pass to (reverse) counter B. The carry impulse loads the content of the latch to counter B. The latch is connected with the reverse counter such that the number written to this counter is $2M$ times smaller than the number introduced to the latch. This can be readily achieved by omitting M most significant bites of counter B. Because the number loaded to counter B is $2M$ times smaller than the number in the latch, the carry impulses of counter B have frequency $2M$ times greater than the frequency of the impulses at the output of the differentiator. The carry impulses are fed to a D flip-flop, which divides their frequency by two. In this way, the output frequency is $2M$ greater than input frequency f_o as long as the frequency of impulse generator f_g is much greater than $2Mf_o$.

33

Function and Signal Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

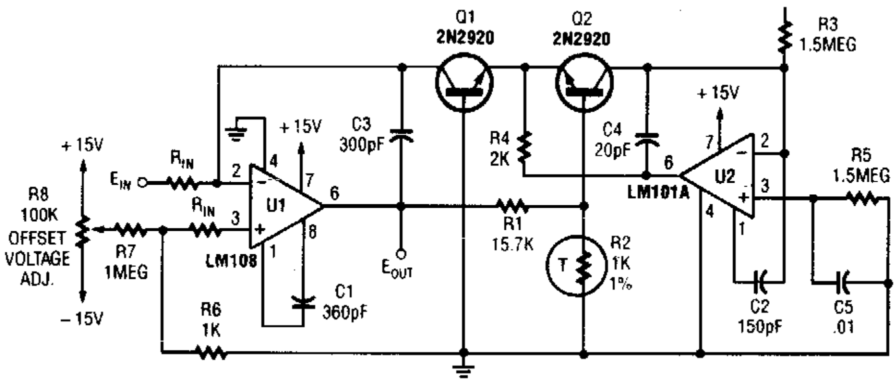
FUNCTION GENERATOR



ELECTRONIC DESIGN *FIG. 33-1*

This function generator, based on an LT1016 high-speed comparator, will generate from a single +5-V supply. The slow rate of the op amps used determines the maximum useable frequency of this circuit.

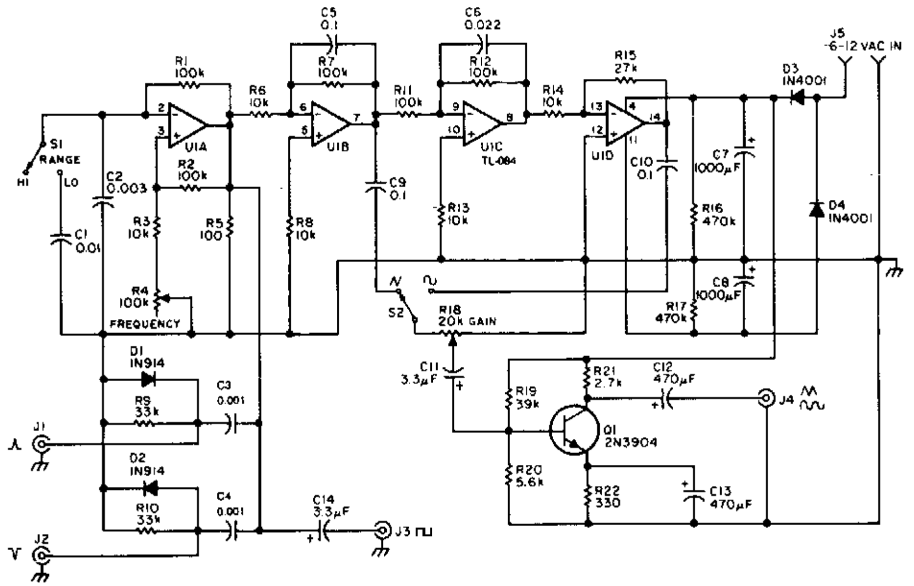
100-dB DYNAMIC-RANGE LOG GENERATOR



POPULAR ELECTRONICS *FIG. 33-2*

E_{OUT} = constant × (Log E_{IN}). This circuit has 100-dB dynamic range, which is five decades of voltage change at the input.

FUNCTION GENERATOR

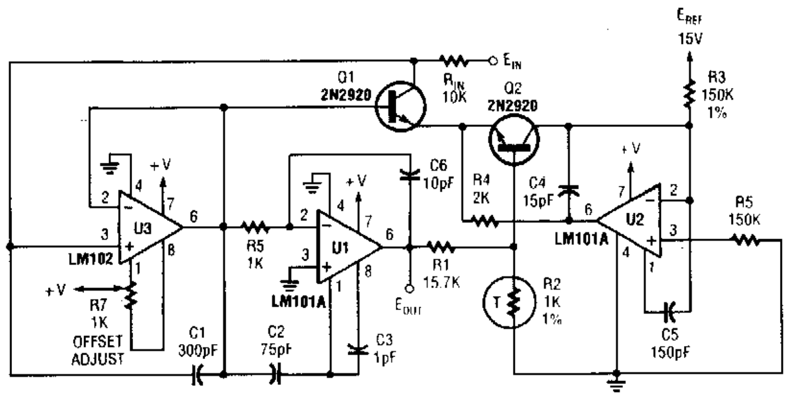


73 AMATEUR RADIO TODAY

FIG. 33-3

A quad op amp makes up the heart of this function generator. U1-a generates a square wave, and outputs this to J3. J1 and J2 are pulse outputs obtained by differentiating the square wave. Integrator U1-b generates a triangle-wave shaper to obtain a sine wave. Q1 is an output amplifier.

FAST LOGARITHM GENERATOR

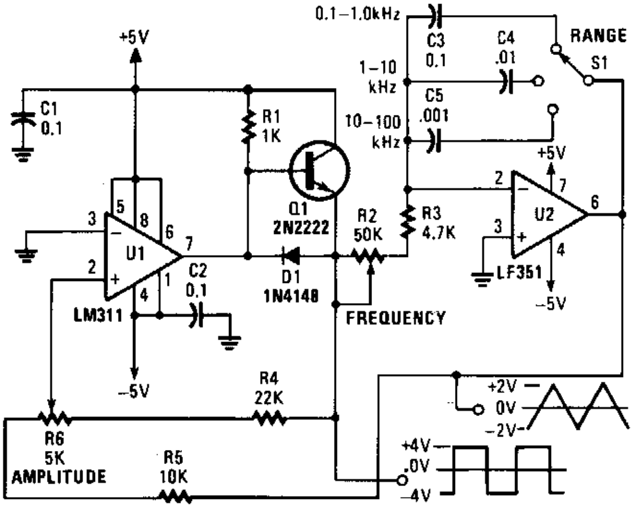


POPULAR ELECTRONICS

FIG. 33-4

In this circuit, E_{OUT} = (constant) × log E_{IN}. The circuit should be useable with op amps other than the ones illustrated.

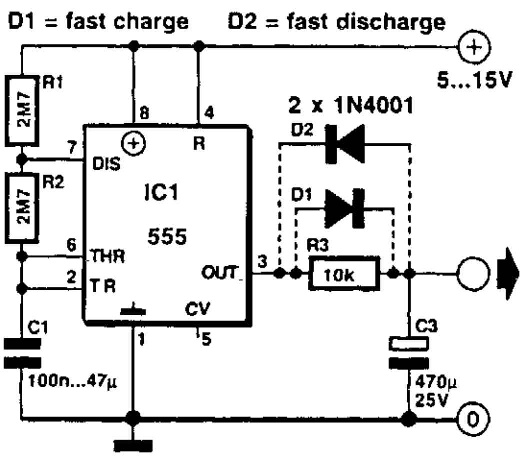
TRIANGLE-WAVE GENERATOR



POPULAR ELECTRONICS *FIG. 33-5*

This is a simple triangle-wave generator using two IC devices and a transistor. The triangle wave is used as feedback to the square-wave generator. S1 allows range switching in three ranges from 100 Hz to 100 kHz. Extra positions could be used to extend the range to lower frequencies, using larger values of capacitance.

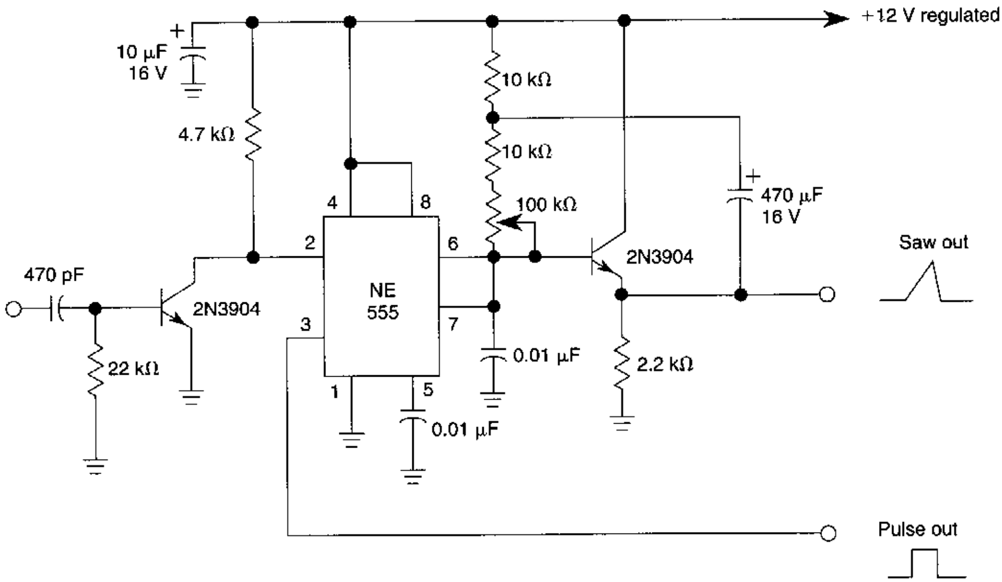
555-BASED RAMP GENERATOR



ELEKTOR ELECTRONICS *FIG. 33-6*

This circuit is used to generate a ramp voltage for tuning a radio receiver. An NE555, running at about 0.1 Hz, is used as an astable multivibrator.

TRIGGERED SAWTOOTH GENERATOR

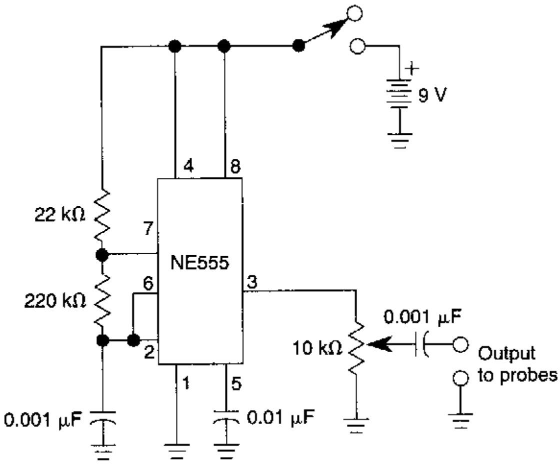


WILLIAM SHEETS

FIG. 33-7

Two 2N3904 transistors and a 555 form a triggered sawtooth generator. A sawtooth or other rising voltage input provides a pulse output when the trigger point is reached.

SIGNAL GENERATOR

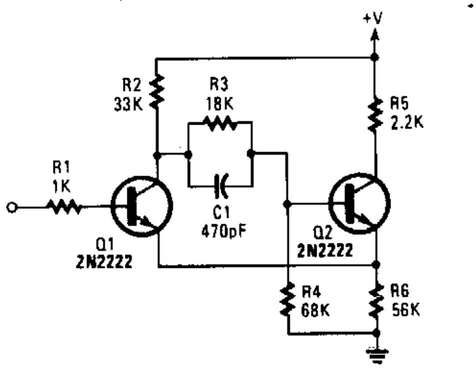


WILLIAM SHEETS

FIG. 33-8

This simple oscillator is rich in harmonics which make this circuit useful for signal tracing applications.

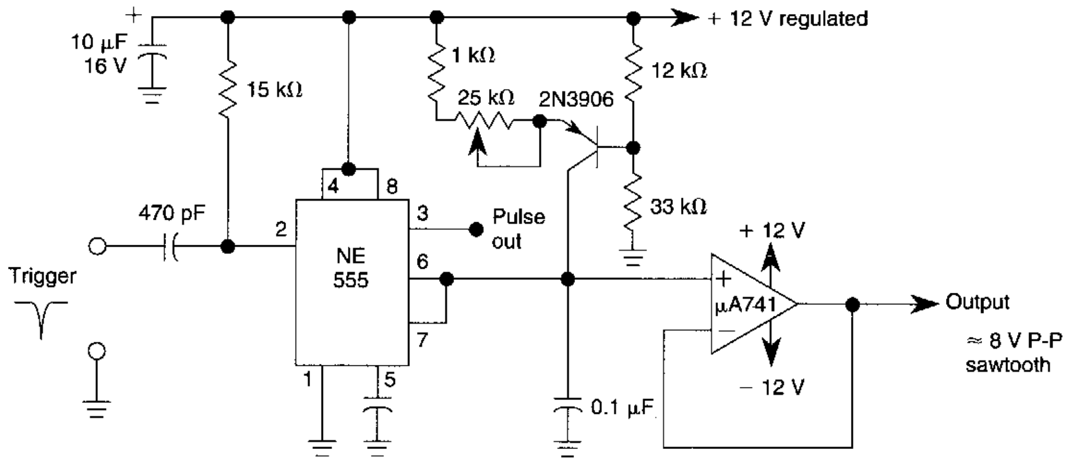
TRANSISTORIZED SCHMITT TRIGGER



POPULAR ELECTRONICS

FIG. 33-9

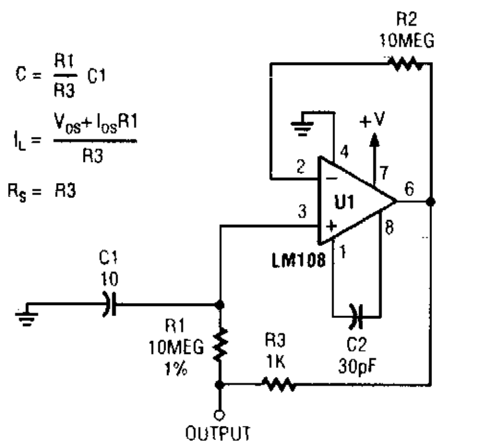
LINEAR SAWTOOTH GENERATOR



WILLIAM SHEETS ***FIG. 33-10***

The 2N3906 transistor is used as a constant-current source, to assure that the 555-based sawtooth generator generates a linear ramp waveform.

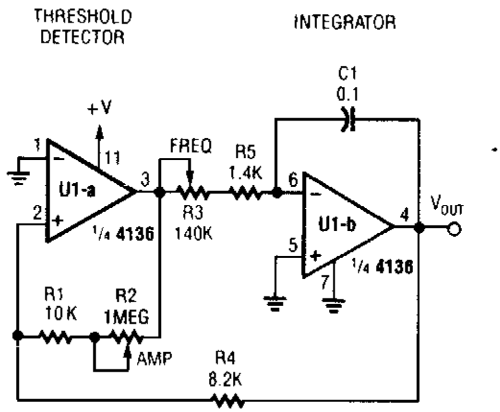
CAPACITANCE MULTIPLIER



POPULAR ELECTRONICS ***FIG. 33-11***

Capacitance multiplier uses the gain of an op amp to produce an effective capacitance—in this case 100,000 μF.

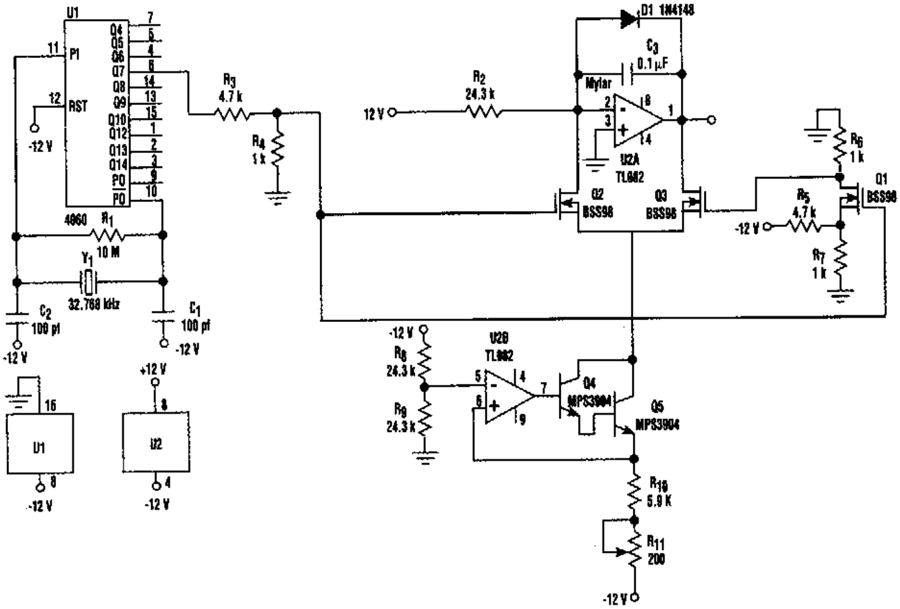
TRIANGLE-WAVE OSCILLATOR



POPULAR ELECTRONICS ***FIG. 33-12***

U1-b acts as an integrator while U1-a is a threshold detector. R2 sets the trip level and therefore the amplitude. R3 controls charging current of C1 and the frequency.

CLOCK-DRIVEN TRIANGLE-WAVE GENERATOR

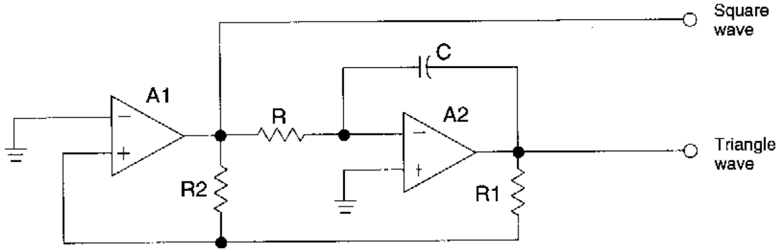


ELECTRONIC DESIGN

FIG. 33-13

U2-a, C3 and R2 operate as an integrator. Q2 and Q3 are alternately switched at 256 cycles. U2-b, Q4, Q5, and R8 through R11 are a constant current generator, and R11 is set for a symmetrical triangular waveform.

TRIANGLE- AND SQUARE-WAVE GENERATOR



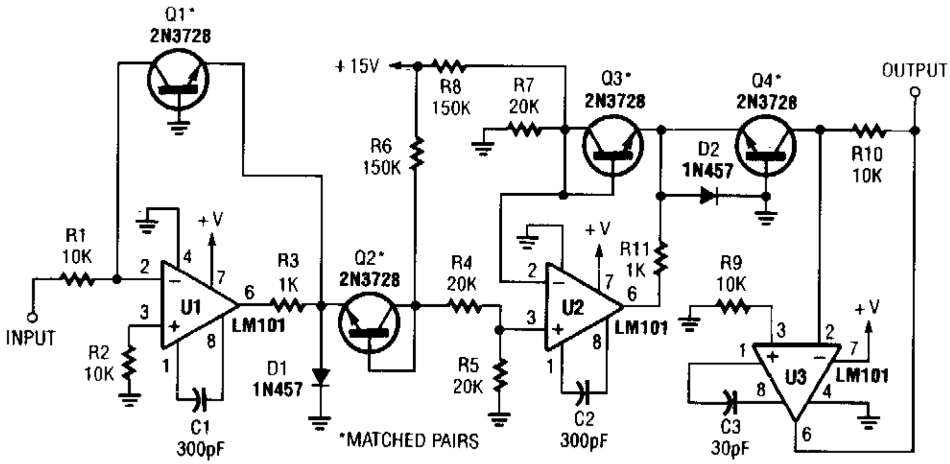
WILLIAM SHEETS

FIG. 33-14

The circuit will generate precision triangle and square waves. The output amplitude of the square wave is set by the output swing of op amp A1, and R_1/R_2 sets the triangle amplitude. The frequency of oscillation in either case is approximately $1/0.69RC$.

The square wave will maintain 50% duty cycle—even if the amplitude of the oscillation is not symmetrical. The use of a fast op amp in this circuit will allow good square waves to be generated to quite high frequencies. Because the amplifier runs open-loop, compensation is not necessary. The triangle-generating amplifier should be a compensated type. A dual op amp, such as the MC1458, can be used for most applications.

ROOT EXTRACTOR



POPULAR ELECTRONICS

FIG. 33-15

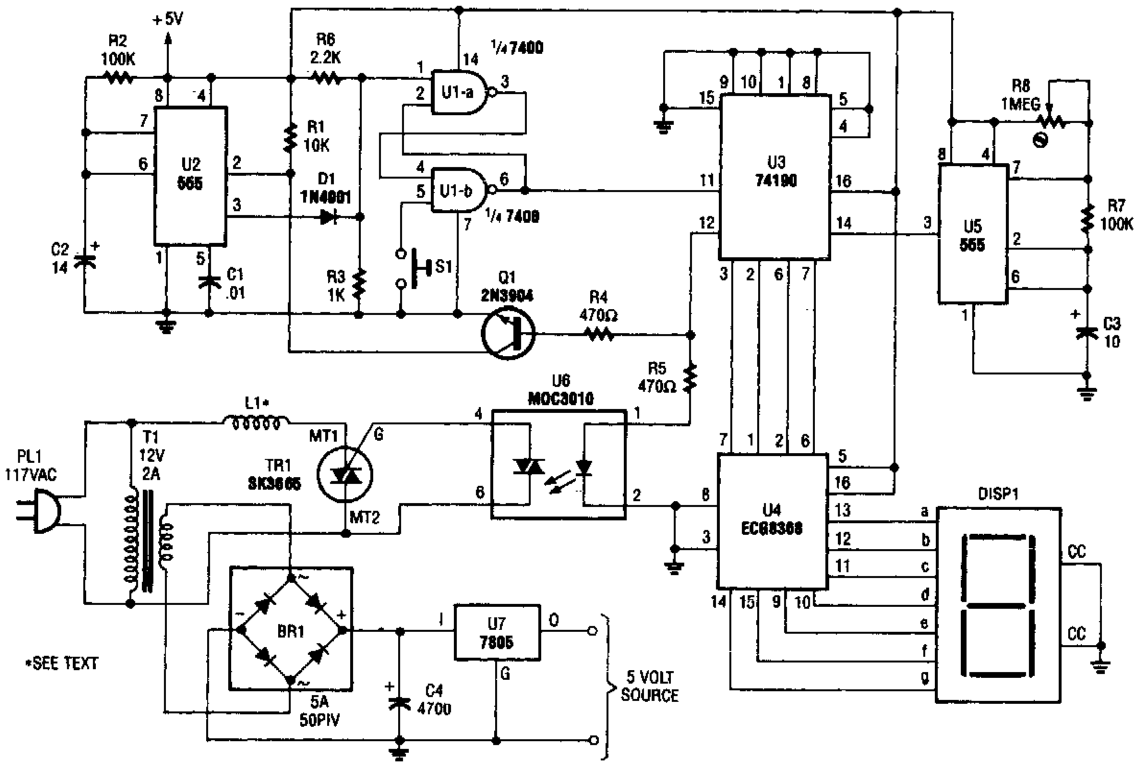
This circuit produces a voltage that is proportional to the root of the input. This gives a logarithmic response, $\log V_{IN}^{N} = N \log V_{IN}$.

34

Game Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTROMAGNETIC RING LAUNCHER

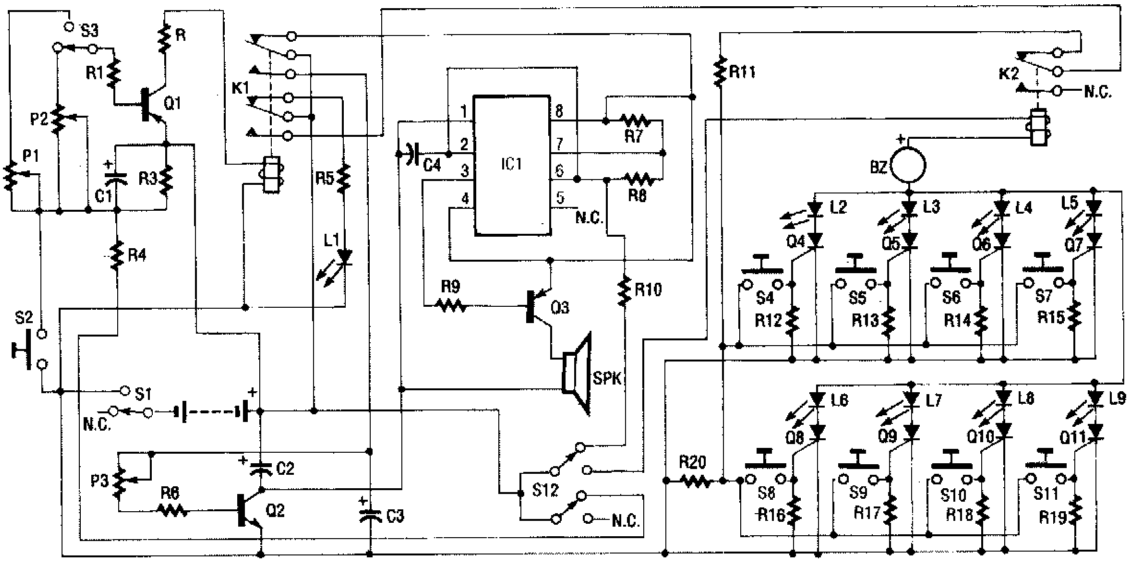


1993 ELECTRONICS HOBBYIST HANDBOOK ***FIG. 34-1***

The electromagnetic ring launcher is comprised of four subcircuits: a clock circuit (built around U5, a 555 oscillator/timer configured for astable operation), a count-down/display circuit (built around U3), a 74190 synchronous up/down counter with BCD outputs that is configured for count-down operation; U4, a ECG8368 BCD-to-7-segment latch/decoder/display driver; and DISP1, a common-cathode seven-segment display), a trigger circuit (comprised of U6), an MOC3010 optoisolator/coupler with Triac-driver output; TR1, an SK3665 200-PIV, 4-A Triac; and a few support components), and a reset circuit (comprised of U1, a 7400 quad 2-input NAND gate; U2, a second 555 oscillator/timer configured for monostable operation; and a few support components).

This circuit is that of a repulsion coil (L1) used to demonstrate the principle of electromagnetic repulsion by propelling a metal ring around the core of L1 through the air. A countdown circuit is provided to count seconds before launch.

QUIZ MASTER



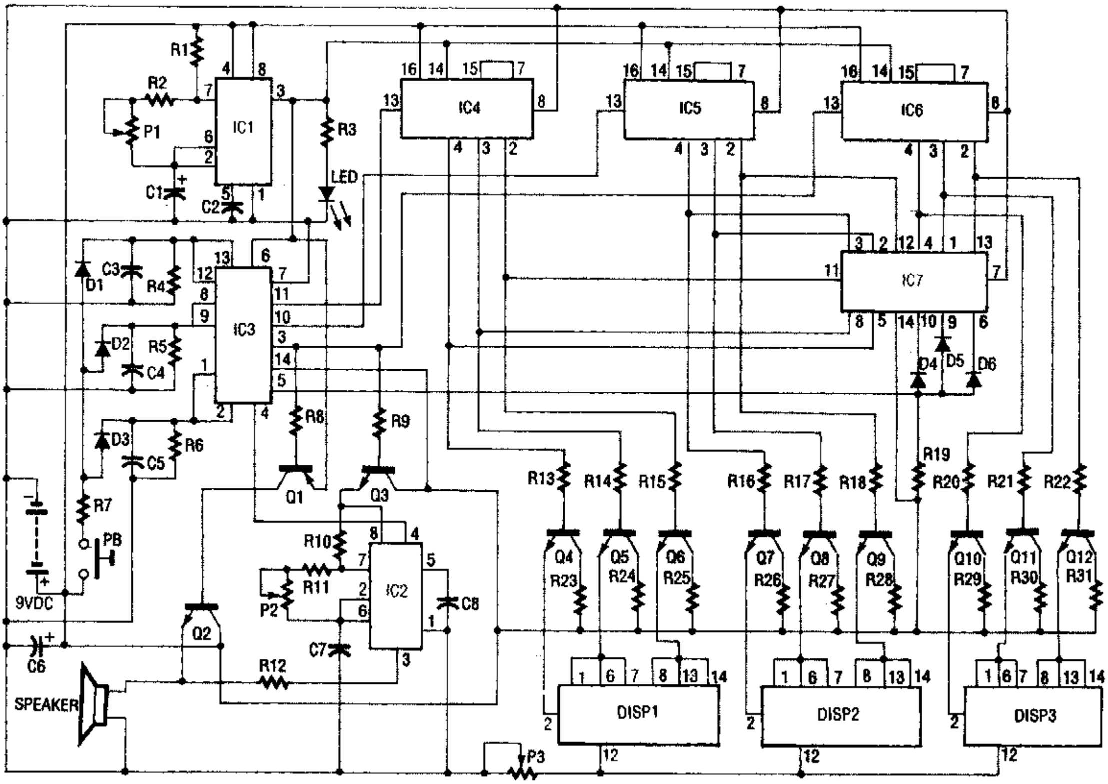
1991 PE HOBBYIST HANDBOOK

FIG. 34-2

Up to eight players each have their own answer button to press, corresponding to the four Red Team and four Green Team LEDs on the master control board. As soon as the first contestant who thinks that he knows the answer presses the button, a loud tone sounds, all other contestants are locked out, and the contestant's indicator LED lights on the control board so that it's obvious who buzzed in first.

The control board also features two selectable "time out" periods—each adjustable from 3 to 15 seconds, setting specified time intervals in which the player must answer before the "time's up!" tone sounds. Eight SCRs form the heart of the circuit. The anode of each SCR has a positive (+) bias on it by way of an LED and a negative (–) bias on each cathode. As soon as a contestant depresses his or her switch button (S4 through S11), a positive bias is applied to the respective SCR gate. That bias latches the contestant's SCR on, which in turn lights up the appropriate LED on the master control board. At the same time, the activity of the SCR latching on turns on the answer buzzer (BZ) and locks out all other contestants. The lockout occurs because relay K2 contacts operate to remove the availability of a bias voltage to the gate of the other SCRs.

The other circuitry consists of a timer circuit and a "time's-up" tone-generating circuit. The timer circuit consists of transistor Q1, capacitor C1, resistors R1 through R3, and trimmer resistors P1 and P2. Depending on the adjustment of the trimmer resistors and selection switch S3, a specific time period can be set. The time's-up tone-generating circuit is made up of IC1, transistors Q2 and Q3, and the associated resistors and capacitors. The "on" time of the tone can be set by P3. Relay K1, which is operated by the timer circuit, serves to reset the entire unit for the next question.



1991 PE HOBBYIST HANDBOOK ***FIG. 34-3***

The slot machine's realistic action is provided by seven ICs and three displays, as shown. Two 555 CMOS timer ICs generate pulses. IC1 is used to generate the clock pulses for the entire electronic slot machine. The pulses are coupled from the output (pin 3) to the clock inputs of IC4, IC5, and IC6, the display-driver ICs.

The displays are common-cathode 7-segment LED types. They are wired to display three different symbols, an "L," a "7," and "bar." When all three displays show the same symbols, IC7 (a 4023 triple 3-input NAND gate) decodes a winner and sends a signal to pin 5 of IC3. That IC is a 4001 CMOS NOR gate and it turns on IC2, a 555 timer IC. IC2 actually produces the winner tone on its output, pin 3.

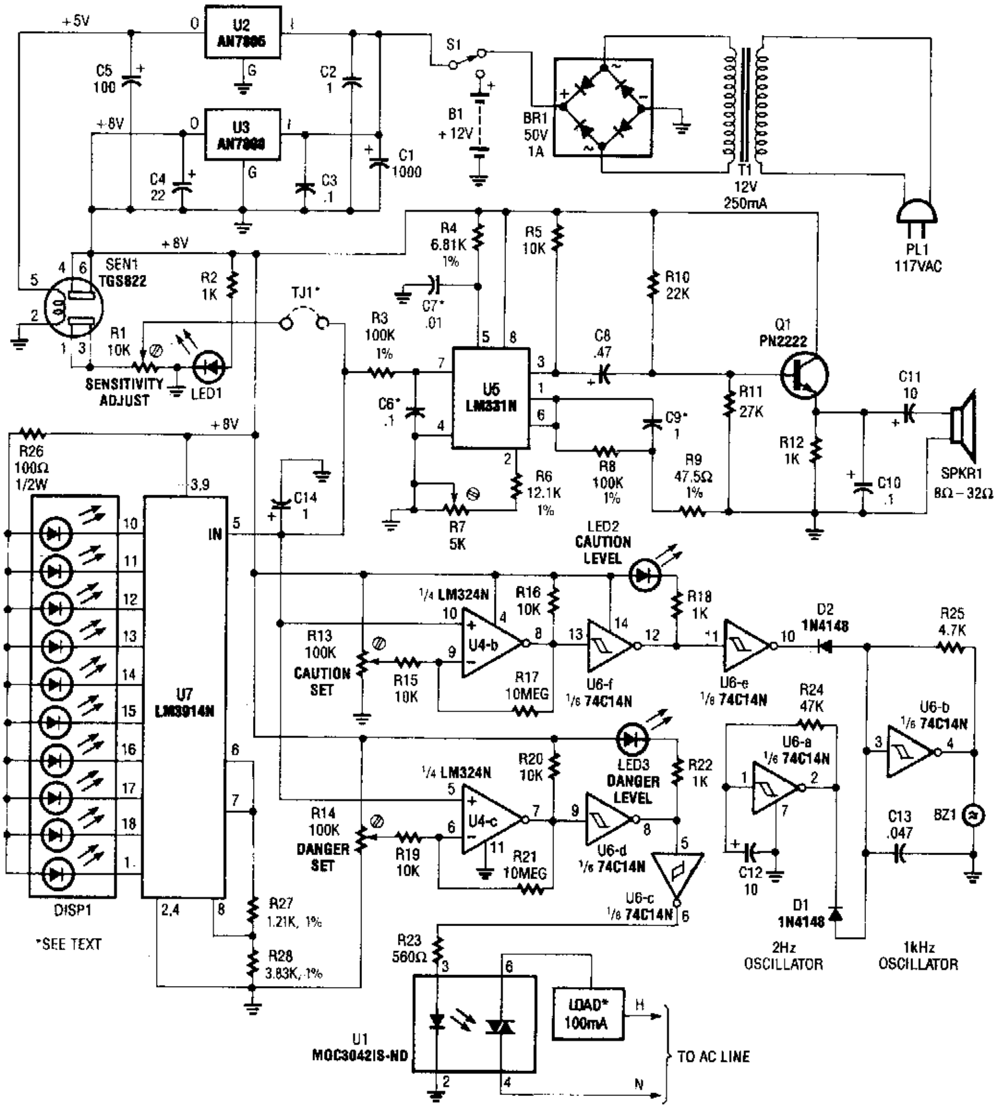
Transistors Q4 through Q12 are used to drive the common-cathode displays. An LED is used to indicate the clock pulses, and a variable resistor is provided for each of these functions. Trimmer resistor P1 controls the overall clock rate, P2 controls the "winner" tone, and P3 controls the display brilliance.

35

Gas Detector Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

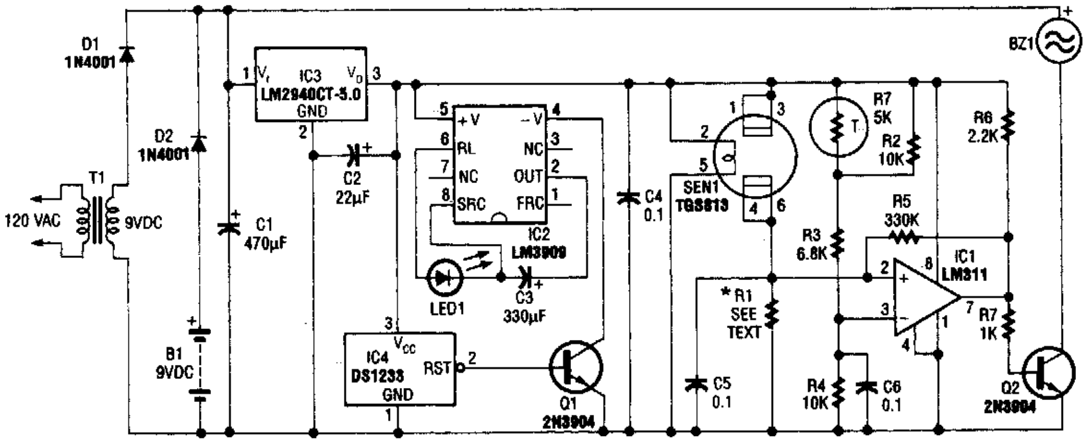
EXPLOSIVE GAS DETECTOR

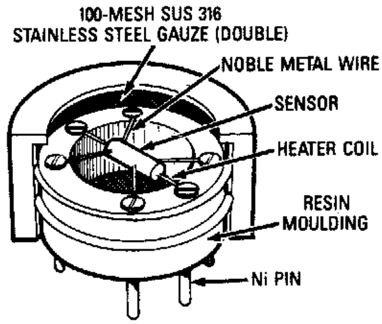


POPULAR ELECTRONICS ***FIG. 35-1***

A gas sensor (TGS823 from Allegro Electronics, Cornwall Bridge, CT 06754) conducts in the presence of explosive gases. U5 is a voltage-to-frequency converter that produces a frequency proportional to the sensor conductance. The output frequency ranges from 100 Hz in clean air to 8 kHz in a contaminated atmosphere. The dc voltage from the sensor also drives bar graph LED U7 and comparators U4-b and U4-c to sense present caution and danger levels. U1 drives an ac load up to 100 mA (relay, indicator, alarm, etc.).

COMBUSTIBLE GAS DETECTOR





THE GAS SENSOR is mainly composed of tin dioxide on a ceramic base; the resistance of the sensor varies depending on the concentration of reducing gases in the air.

ELECTRONICS NOW ***FIG. 35-2***

The circuit shown is useful for the detection of dangerous levels of combustible fumes or gases. It uses a comparator circuit to trigger an alarm buzzer. The sensor's resistant element is connected in series with resistor R1 to form a voltage-divider circuit; R1 is specifically matched to each gas sensor by the manufacturer.

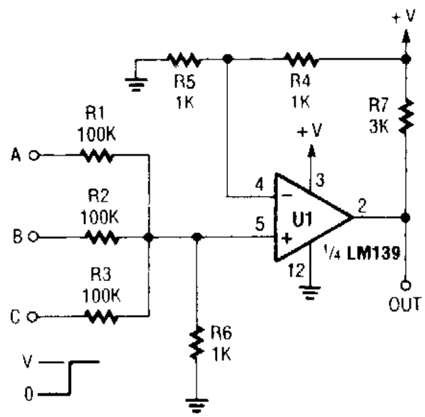
36

Gate Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

AND Gate

AND GATE



POPULAR ELECTRONICS

FIG. 36-1

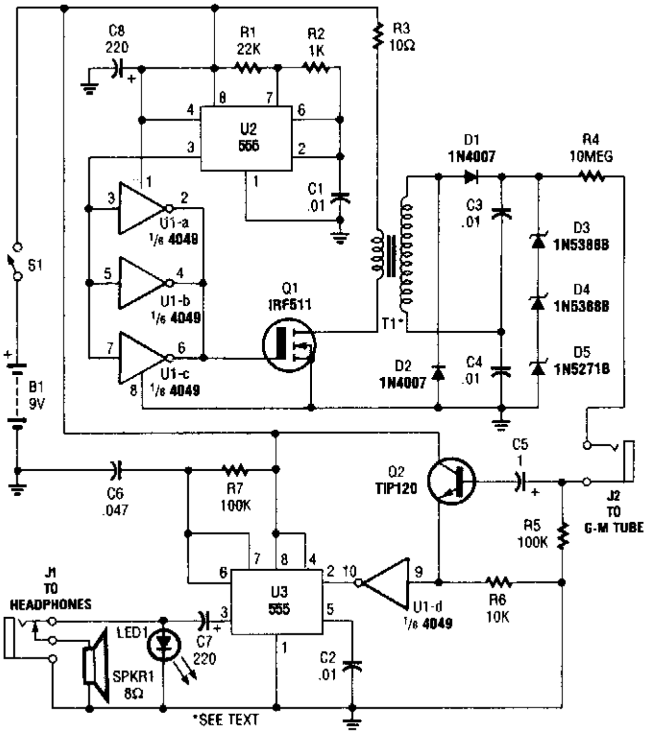
A left-over section of a quad op amp can be used to save cost and eliminate an extra logic chip for this AND gate.

37

Geiger Counter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

GEIGER COUNTER I



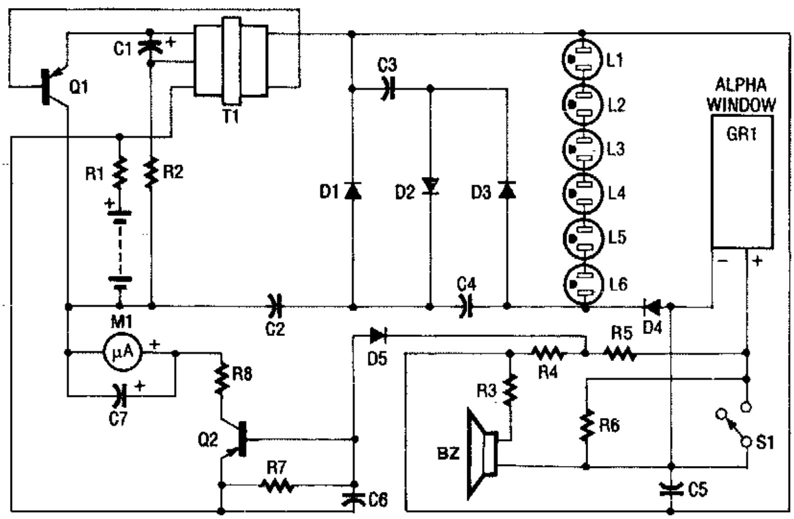
POPULAR ELECTRONICS *FIG. 37-1*

The circuit is built around a 4049 hex inverter (U1), a pair of 555 oscillator/timers (U2 and U3), two transistors, a Geiger-Muller tube, and a few additional support components. The first 555 (U2) is configured for astable operation. The output of U2 (a series of negative-going pulses) at pin 3 is fed to three parallel-connected inverters (U1-a, U1-b, and U1-c). The positive-going output pulses of the inverters are fed to the gate of Q1, causing it to toggle on and off.

The output of Q1, which is connected in series with the primary of step-up transformer T1, produces a stepped-up series of pulses in T1's secondary. The output of T1 (approximately 300 V) is fed through a voltage doubler (consisting of D1, D2, C3, and C4), producing a voltage of around 600 V. Three series-connected Zener diodes (D3, D4, and D5) are placed across the output of the voltage doubler to regulate the output to 500 V, fed through R4 (a 10-MΩ current-limiting resistor) and J2 to the anode of the GM tube. The limiting resistor also allows the detection ionization to be quenched.

The cathode side of the tube is connected to ground through a 100-kΩ resistor, R5. When a particle is detected by the GM tube, the gases within the tube ionize, producing a pulse across R5. That pulse is also fed through C5 and applied to the base of Q2 (a TIP120 npn transistor), where it is amplified and clamped to 9 V. The output of Q2 is inverted by gate U1-d, then it is used to trigger U3 (the second 555, which is configured for monostable operation). The output of U3 at pin 3 causes LED1 to flash, and produces a click that can be heard through speaker SPKR1 or headphones. The circuit is powered by a 9-V alkaline battery and draws about 28 mA when not detecting radiation.

GEIGER COUNTER II



BZ ....Blue Piezo Buzzer
C1 .... 4.6-µF Electrolytic Capacitor
C2–C4 ....0.005-µF 1-kV Disc Capacitor
C5 .... .01-µF 1-kV Disc Capacitor (103 M)
C6 .... .1-µF 100-V Mylar Capacitor (104 k)
C7 .... 33-µF Electrolytic Capacitor
D1–D5 . . .1N4007 Diodes
GR1 .... Alpha Window Geiger Mueller Tube
L1–L6 .....Neon Lamps
M1 .... 0–200 Microamp Meter
Q1 .... 02-GE PNP Power Transistor
Q2 .... 2N3906 Transistor
R1 .... 47-ohm Resistor
R2, R3 ... 3.9-k Resistor
R4, R5 ... 4.7-Meg Resistor
R6 .... 220-k Resistor
R7 .... 27-k Resistor
R8 .... 18-kΩ Resistor
S1 .... SPDT Slide Switch
T1 .... Inverter Transformer

PE HOBBYIST HANDBOOK ***FIG. 37-2***

Q1 is a pnp power transistor used in conjunction with a ferrite transformer to form a blocking-type oscillator. This oscillator is a fixed-frequency type, and the feedback to sustain oscillations is from capacitor C1. Because of the turns ratio of T1, the small ac voltage produced on its primary is converted to a large ac voltage on its secondary. That high-voltage ac is applied to the voltage tripper stage, which consists of capacitors C2, C3, and C4 and diodes D1, D2, and D3. The resultant voltage is now over 800 V and it is regulated by neon lamps L1 through L6. Diode D4 rectifies the high voltage and applies it to the cathode lead of the GM tube. The positive (+) bias on the GM tube is applied to the anode by way of load resistors R4 and R5. Each time a radioactive particle strikes the GM tube, it causes the gas inside to ionize. This ionization of the gas creates a pulse, which drives the piezo speaker and is also coupled by diode D5 to the base of Q2. Transistor Q2 is a pnp type and is used to "integrate" the pulses in conjunction with capacitor C6. That produces a dc voltage level, which is in proportion to the quantity of pulses arriving at the base of Q2. The collector of Q2 is connected through resistor R8 to the (+) terminal of the meter. The other side of the meter goes directly to (–) of the battery.

38

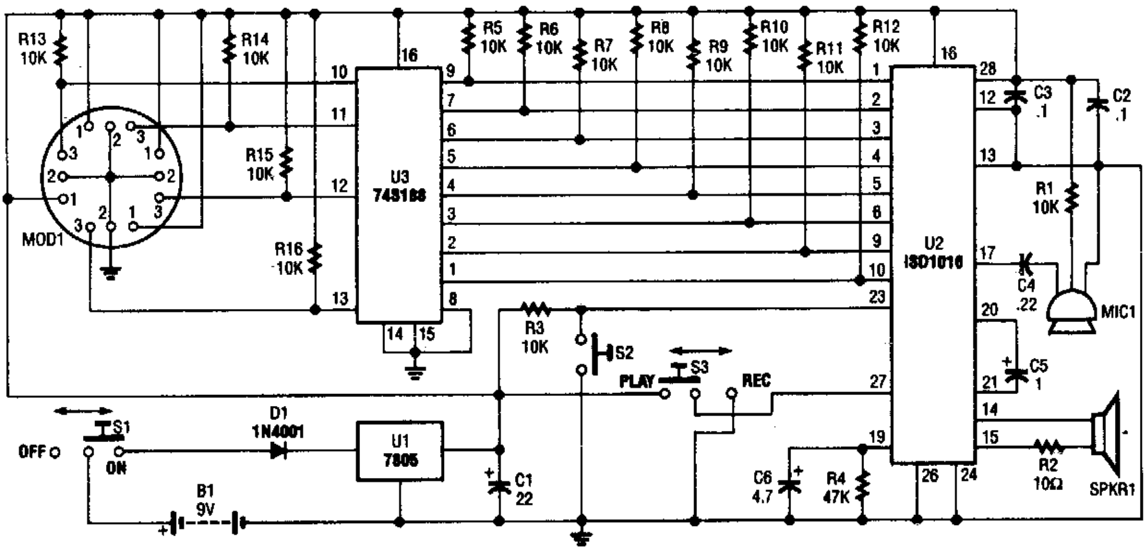
Hall Effect Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

THE TALKING COMPASS

TABLE 1—74S188 TRUTH TABLE

Directory	Input					Output								Decimal Equivalent
	A4	A3	A2	A1	A0	B0	B1	B2	B3	B4	B5	B6	B7	
North	L	H	L	H	H	0	0	0	0	0	0	0	1	1
N.W.	L	L	L	H	H	0	0	0	1	0	1	0	0	20
West	L	L	H	H	H	0	0	1	0	1	0	0	0	40
S.W.	L	L	H	H	L	0	0	1	1	1	1	0	0	60
South	L	H	H	H	L	0	1	0	1	0	0	0	0	80
S.E.	L	H	H	L	L	0	1	1	0	0	1	0	0	100
East	L	H	H	L	H	0	1	1	1	1	0	0	0	120
N.E.	L	H	L	L	H	1	0	0	0	1	1	0	0	140



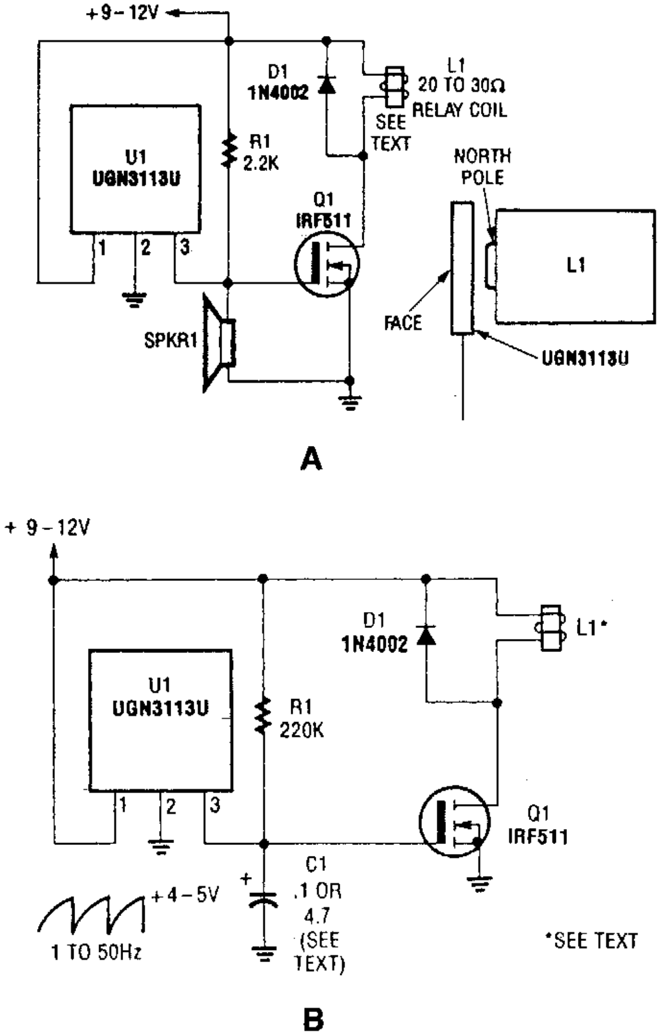
1993 ELECTRONICS HOBBYISTS HANDBOOK

FIG. 38-1

A talking compass is made up using a Hall-effect direction sensor (MOD1) and an ISD1016 analog audio storage device. It is possible to program eight two-second announcements, for each of the eight main compass directions.

The Talking Compass is comprised of a digital compass (MOD1), and ISD1016 analog storage device (U2), a 74S188 preprogrammed PROM (U3), and a handful of additional components.

UNUSUAL HALL-EFFECT OSCILLATORS



POPULAR ELECTRONICS

FIG. 38-2

Although not intended for this application, Hall-effect switch can be used as the basis for a rather unusual oscillator. The oscillator can be reconfigured, as shown in Fig. B, to allow the circuit's oscillating frequency to be controlled via an RC network, comprised of R1 and C1.

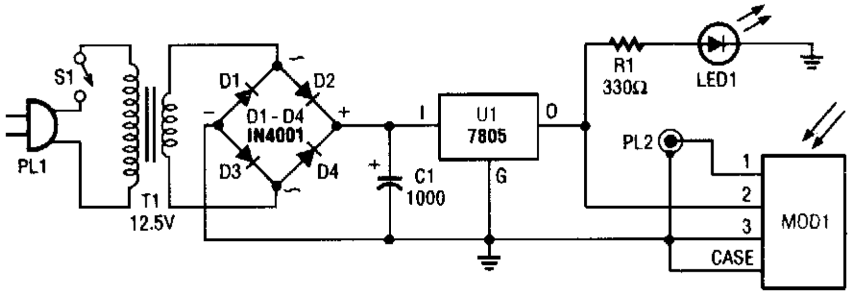
39

Infrared Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Remote-Control Analyzer
IR-Pulse-to-Audio Converter
IR-Controlled Remote A/B Switch
Simple IR Detector
Infrared Receiver
Selective Preamplifier for Infrared Photodiode
Wireless IR Headphone Transmitter
Wireless IR Headphone Receiver
Infrared Remote-Control Tester
Pulsed Infrared Transmitter for On/Off Control
Very Simple IR Remote-Control Circuit
IR Receiver
Remote-Control Tester

REMOTE-CONTROL ANALYZER



POPULAR ELECTRONICS *FIG. 39-1*

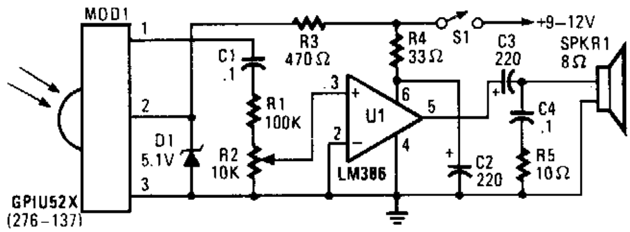
A schematic diagram for the remote analyzer is shown. The circuit is powered from a simple 5-V supply, consisting of PL1, S1, T1, a bridge rectifier (comprised of D1 through D4), capacitor C1, and a common 5-V regulator, U1. Switch S1 is the on/off control and is optional. The power-supply transformer used in the prototype is a 12.6-Vac unit, but any transformer that can supply at least 5.6-Vac will do. The 12.6-V unit was used solely because of its availability.

The output of T1 is full-wave rectified by diodes D1 through D4 and filtered by C1. The bumpy dc output from the capacitor is regulated down to 5 V by U1, a 7805 integrated regulator. LED1 acts as a power indicator to let you know that the circuit is active.

The 5-Vdc powers a GPIU52X infrared-detector module* (MOD1), which demodulates the 40-kHz carrier used by most infrared remotes. After demodulation, the resulting logic pulses are sent to an oscilloscope via PL2, a BNC connector.

*Radio Shack part #276-137

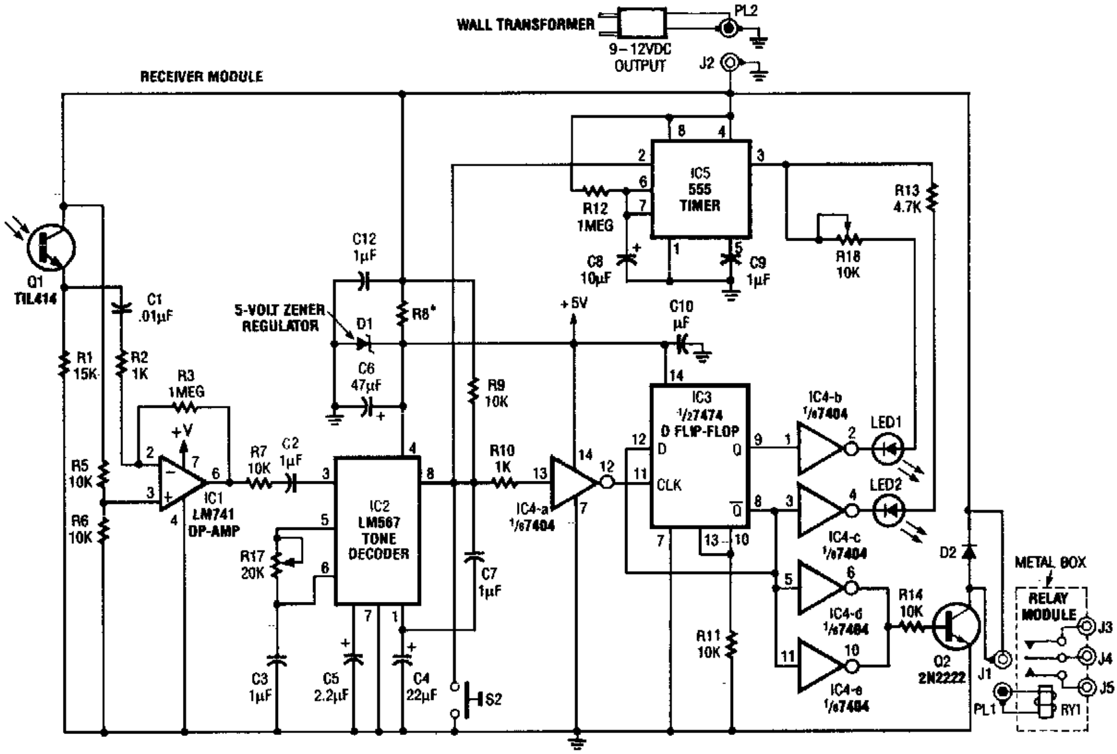
IR-PULSE-TO-AUDIO CONVERTER



POPULAR ELECTRONICS *FIG. 39-2*

If your ear is good, you can use this IR-pulse-to-audio converter to troubleshoot infrared remote-controls. It is also a good project for detecting infrared-light sources. A photo cell module (Radio Shack P/N 276-137) detects IR radiation and drives audio IC U1. This circuit is useful for troubleshooting IR remote controls.

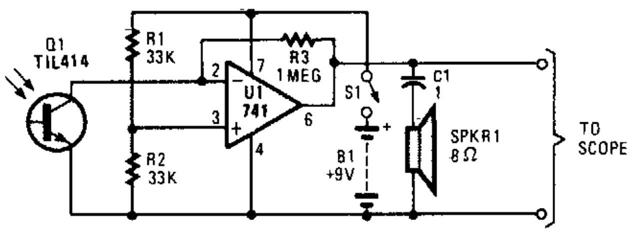
IR-CONTROLLED REMOTE A/B SWITCH



RADIO ELECTRONICS ***FIG. 39-3***

Useful for A/B control, the IR receiver shown controls a relay from an infrared beam that has a pulsed tone-modulated signal. Q1 is the photo receptor feeding op amp IC1, tone decoder IC2, and flip-flop IC3. IC5 turns off the indicator LEDs after about 15 seconds.

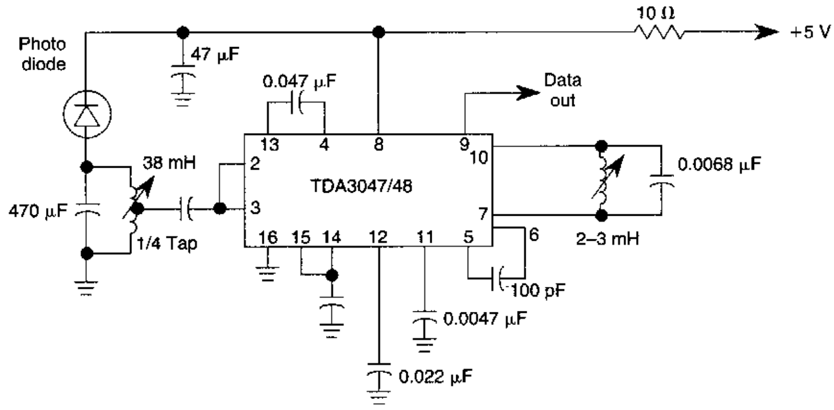
SIMPLE IR DETECTOR



POPULAR ELECTRONICS ***FIG. 39-4***

Useful for IR detection, this circuit uses an op amp of the 741 family (or similar) to detect and amplify IR pulses.

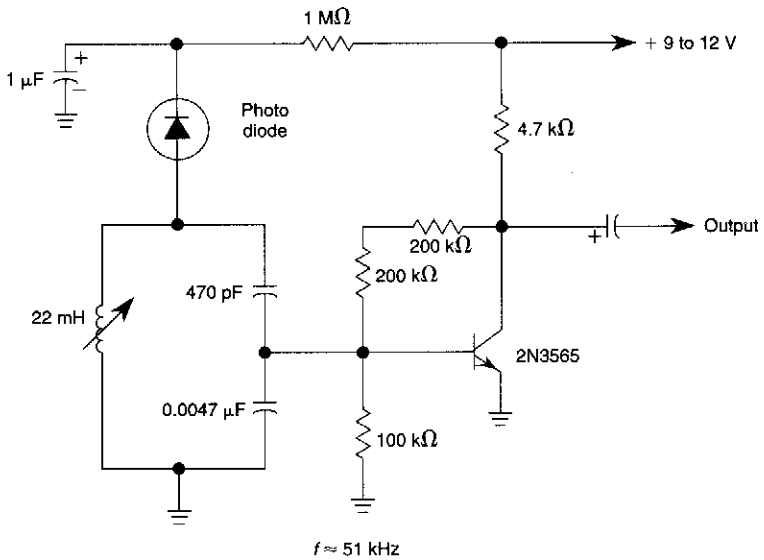
INFRARED RECEIVER



WILLIAM SHEETS *FIG. 39-5*

The circuit operates from a 5-V supply and has a current consumption of 2 mA. The output is a current source that drives or suppresses a current of more than 75 µA with a voltage swing of 4.5 V. The *Q*-killer circuit eliminates distortion of the output pulses because of the decay of the tuned input circuit at high input voltages. The input circuit is protected against signals of more than 600 mV by an input limiter. The typical input is an AM signal at a frequency of 36 kHz.

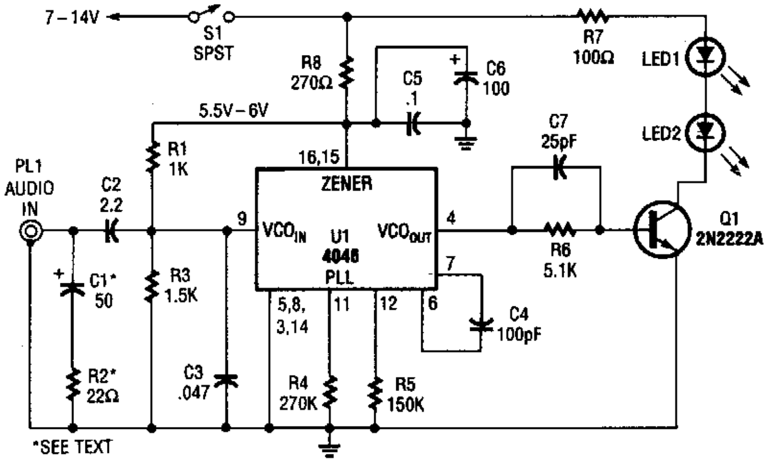
SELECTIVE PREAMPLIFIER FOR INFRARED PHOTODIODE



WILLIAM SHEETS *FIG. 39-6*

The circuit uses a tuned circuit to achieve frequency selection. Values are for operation at about 51 kHz. The 2N3565 amplifies the output developed by the tuned circuit.

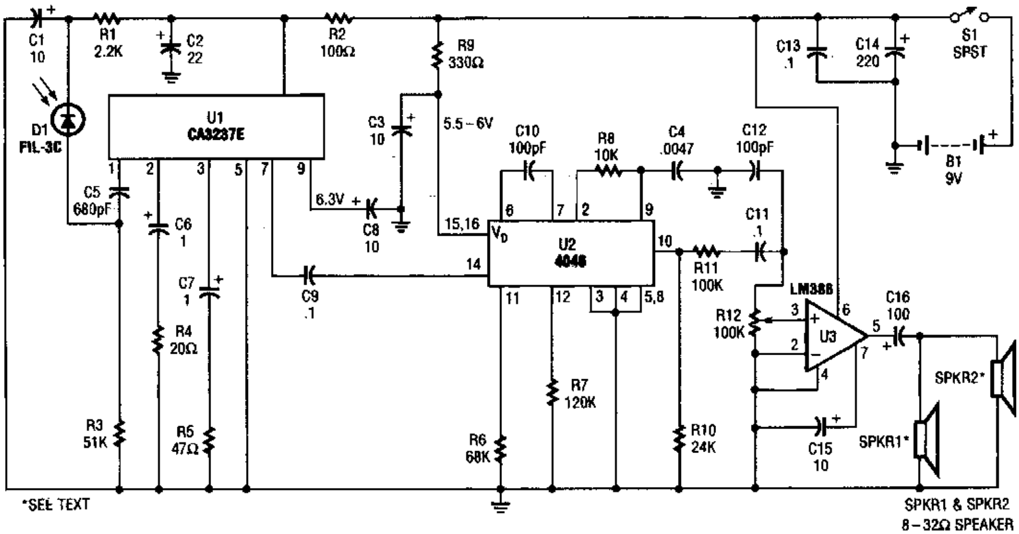
WIRELESS IR HEADPHONE TRANSMITTER



POPULAR ELECTRONICS ***FIG. 39-7***

The transmitter for the wireless headphones is built around a CD4046 CMOS phase-locked loop, coupled with a driver transistor, and a pair of infrared LEDs. Although the CD4046 is comprised of two phase comparators, a voltage-controlled oscillator (or VCO), a source follower, and a zener reference, only its VCO is used in this application.

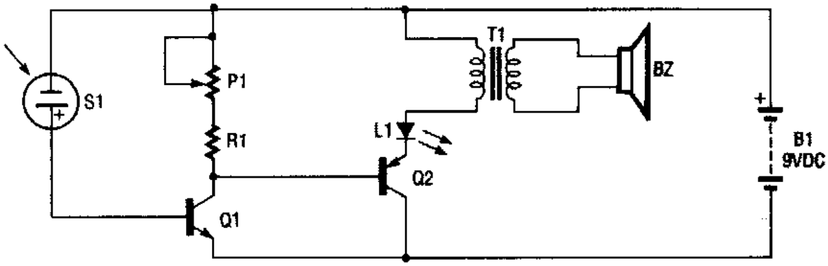
WIRELESS IR HEADPHONE RECEIVER



POPULAR ELECTRONICS ***FIG. 39-8***

IR detector diode D1 intercepts the IR signal at around 40 kHz and feeds it from U1, a high-gain preamp, to PLL, U2, a 4046 configured to serve as an FM detector. U3 is an audio amplifier that feeds a pair of headphones or a speaker.

INFRARED REMOTE-CONTROL TESTER

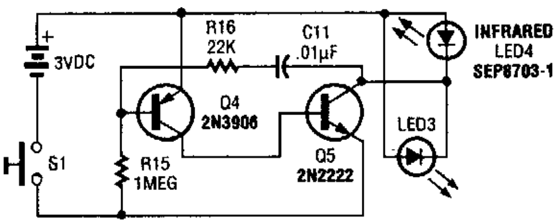


1991 PE HOBBYIST HANDBOOK *FIG. 39-9*

The infrared remote-control tester uses a sensitive PN-type solar sensor that is connected directly to a Darlington amplifier made up of transistors Q1 and Q2. Biasing is provided by R1 and P1, a variable resistor that serves as a sensitivity control. The collector lead of Q1 is the output lead of the Darlington amp, and it is connected to a red LED and the primary of transformer T1. The function of T1 is to convert the low-voltage output signal to a level high enough to drive a small piezo disc. That disc makes a clicking sound when the sensor picks up an infrared signal that is varying in frequency or amplitude. The infrared sensor will also pick up visible light. The use of an IR filter (Wratton #87) is recommended.

BZ	Piezo Disc
L1	Jumbo Red LED
P1	2-MΩ Trimmer Resistor
Q1	2N3904 Transistor
Q2	2N3906 Transistor
R1	270-Ω Resistor
S1	Solar Sensor
T1	Audio Transformer

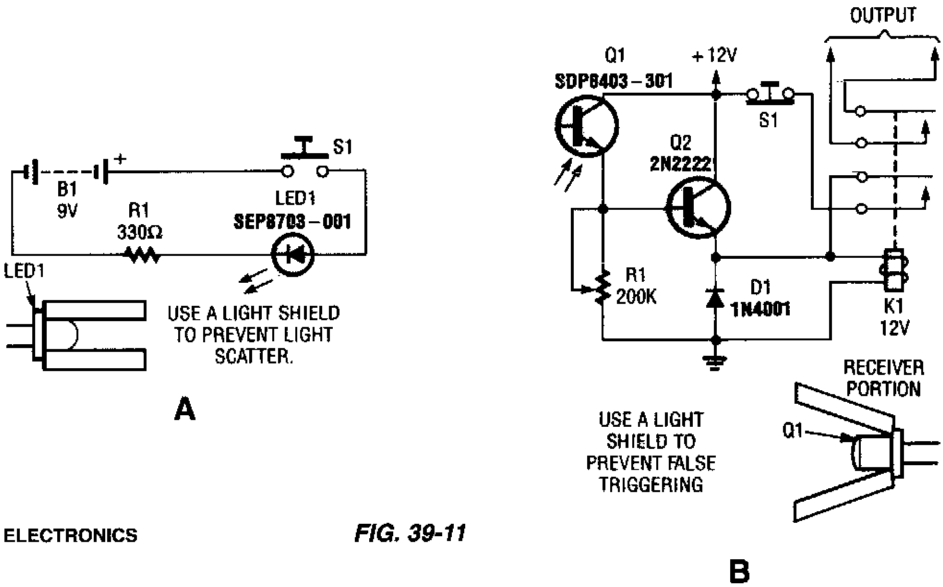
PULSED INFRARED TRANSMITTER FOR ON/OFF CONTROL



RADIO ELECTRONICS *FIG. 39-10*

This transmitter consists of an oscillator and LEDs. It generates a pulsed tone of around 850 Hz.

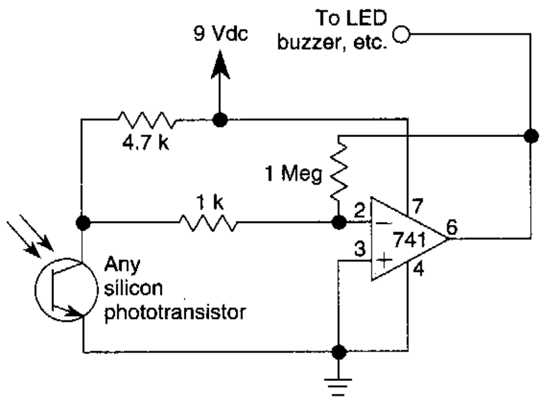
VERY SIMPLE IR REMOTE-CONTROL CIRCUIT



POPULAR ELECTRONICS *FIG. 39-11*

Here is a complete IR remote-control system that consists of a simple transmitter (A) and an equally simple receiver (B).

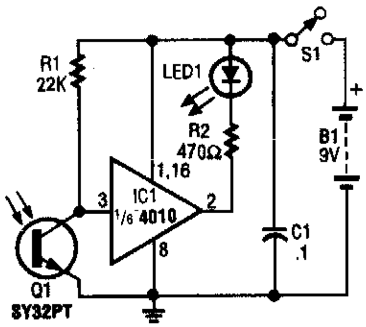
IR RECEIVER



ELECTRONICS NOW *FIG. 39-12*

This circuit is just about the simplest IR receiver you can build. The parts are cheap, the layout is not critical, and a 9-V battery will last a long time.

REMOTE-CONTROL TESTER



ELECTRONICS NOW *FIG. 39-13*

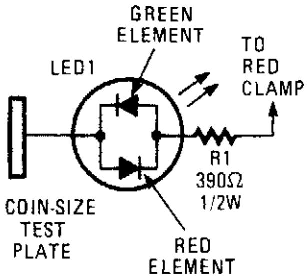
The IR Tester circuit lets you know if the button you press on a remote control is working. Q1 is a photo transistor that is activated by IR energy.

40

Indicator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

POLARITY INDICATOR



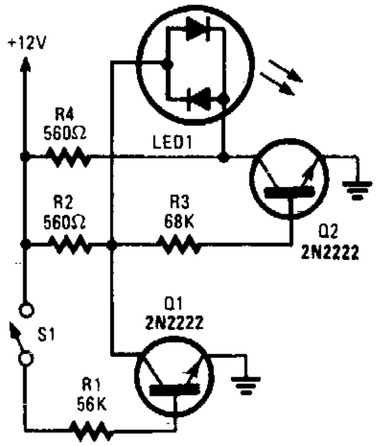
POPULAR ELECTRONICS **FIG. 40-1**

This circuit consists of a tri-color LED, a resistor, wire, and a coin-size test plate. You will have to build two such circuits—one for each black clamp on a set of auto battery jumper cables. The author installed the circuits inside the black clamps themselves using lengths of wire to make the connections to the red clamps.

The first step is to connect one red clamp to what you believe is the positive post on the okay battery. Then, touch the test plate on the black clamp at the end of the cable to the negative terminal on the good battery. The LED will light red if the red clamp is on the wrong terminal. If so move the clamp to the other post and check again. If all is well, the LED will light green. Pick up the other black clamp and connect it to the remaining post on the good battery.

Connect the remaining red clamp to what you assume to be the positive terminal on the bad battery. Now, touch the test plate on the remaining clamp to the engine block or a bare area on the dead car's frame. If the LED appears or doesn't glow, switch the red clamp to the other terminal and test again. When the LED glows green, attach the black clamp to the car's frame (which will prevent any sparks from occurring near the battery). When you remove the clamps, take the clamps off in reverse order to avoid sparks.

BI-COLOR INDICATOR



POPULAR ELECTRONICS *FIG. 40-2*

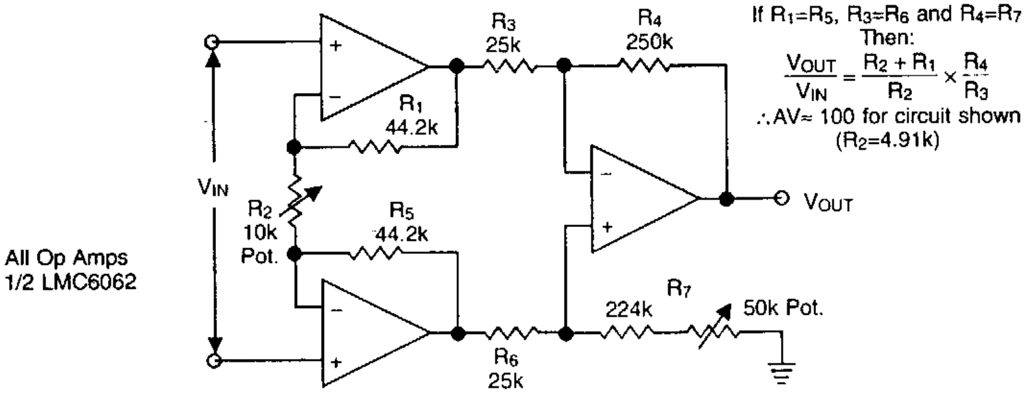
With S1 open, base bias is supplied to Q2 through a voltage divider (formed by R2 and R3), thus turning on the green element in the LED. That indicates that power is being supplied to the project. If you close S1, current through R1 biases Q1 on, thereby grounding the voltage divider and turning off Q2. That reverses the flow of current through the LED, which causes its red element to light. That indicates that the circuit is under power and S1 (really a DPDT switch), whose remaining section controls another circuit, is active. In this circuit, a bi-color LED is used to indicate when a circuit is under power and the status of S1. In that way, the LED does the job of two indicators.

41

Instrumentation Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

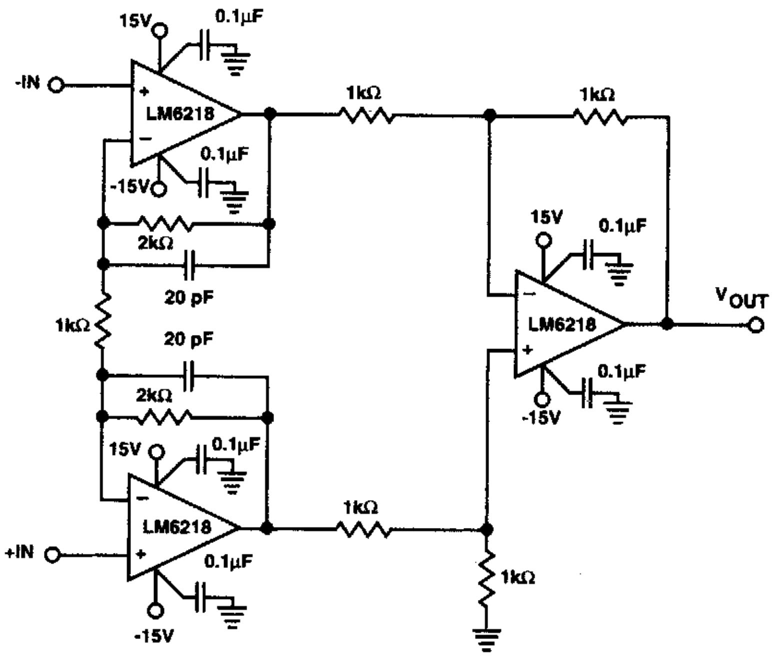
LMC6062 INSTRUMENTATION AMPLIFIER



NATIONAL SEMICONDUCTOR *FIG. 41-1*

Useful for +5-V single-supply applications, this op amp circuit features low drain (around 1 mA), high input resistance (10^{14} Ω), and low bias current ($\approx 10^{-14}$A).

LM6218 HIGH-SPEED INSTRUMENTATION AMPLIFIER



NATIONAL SEMICONDUCTOR

FIG. 41-2

This amplifier features 400-µsec settling time (to 0.01%), 140-V/µsec slow rate, and 17-MHz gain-bandwidth product. The supply voltage can be ±5 to ±20 V.

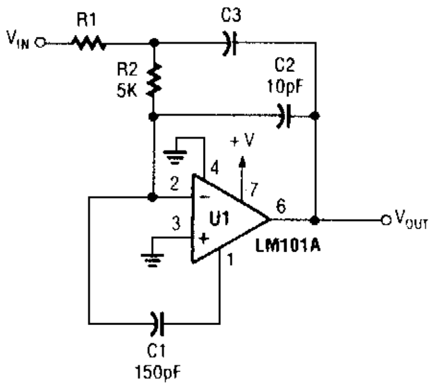
42

Integrator Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Fast Integrator

FAST INTEGRATOR



V_{OUT} is the integral of $V1$ in this circuit.

$$\frac{V_{OUT}}{V_{IN}} \approx \frac{1}{C_3} \frac{V_{IN}(A)}{R} \text{ dt.}$$

POPULAR ELECTRONICS

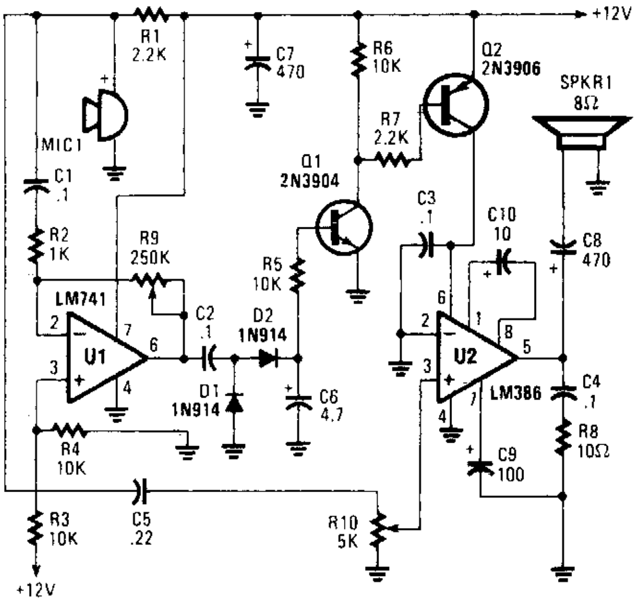
FIG. 42-1

43

Intercom Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

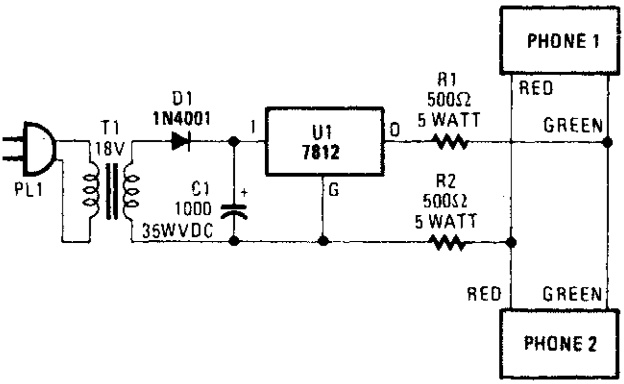
ONE-WAY VOICE-ACTIVATED INTERCOM



POPULAR ELECTRONICS *FIG. 43-1*

An omnidirectional electret microphone can be used to pick up the sound and convert it into an electrical signal. The output of the microphone is fed along two paths. In the first path, the signal is sent to the inverting input at pin 6. In the second path, the microphone signal is fed to the non-inverting input of U2, where it is amplified and output to the speaker, SPKR1.

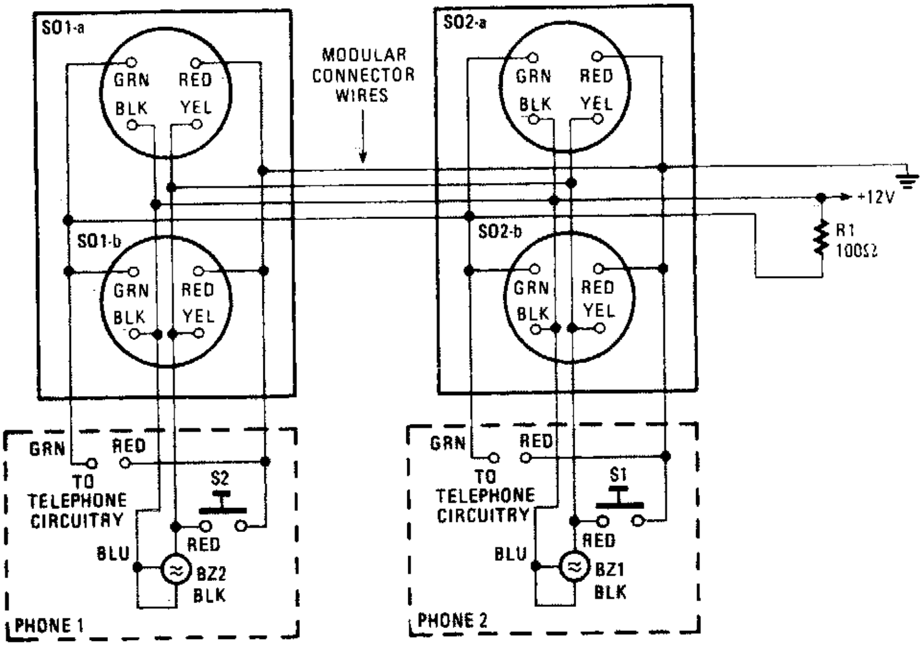
VERY SIMPLE TELEPHONE INTERCOM CIRCUIT



POPULAR ELECTRONICS *FIG. 43-2*

Two telephones can be used as an intercom by using this circuit. Older style rotary phones that are nonelectronic might work best in this application. Also, handsets only might be powered this way.

TELEPHONE INTERCOM



POPULAR ELECTRONICS ***FIG. 43-3***

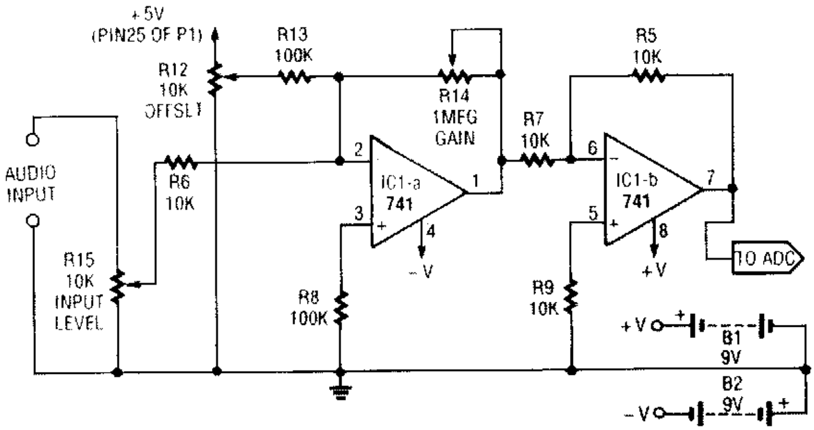
An intercom using dual-modular wall jacks is shown in this circuit. If the wires are available in the home telephone cable, this system can be installed with little trouble.

44

Interface Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

AUDIO-TO-ADC INTERFACE

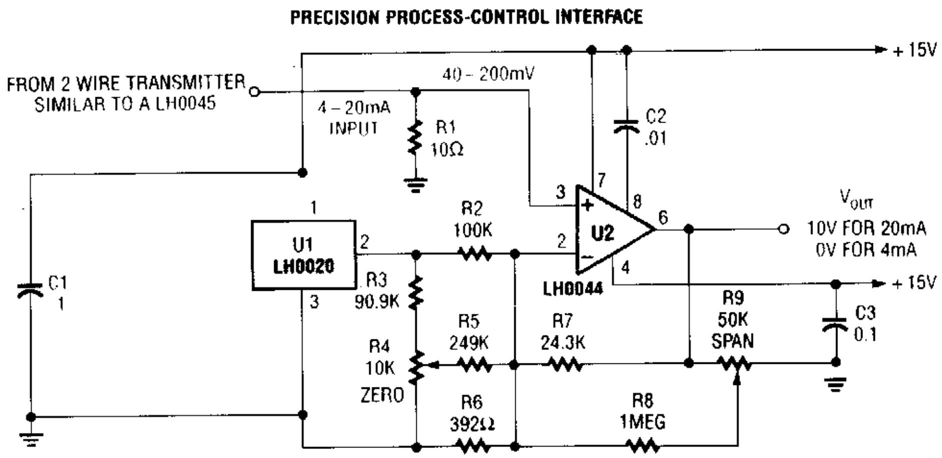


RADIO-ELECTRONICS

FIG. 44-1

This simple general-purpose driver for an analog/digital converter uses two 741 IC devices with adjustable gain and offset. Other op amps might be substituted, but some circuit adjustments might be needed.

PROCESS-CONTROL INTERFACE

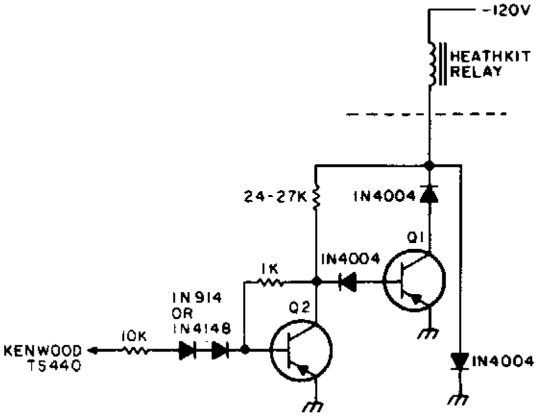


POPULAR ELECTRONICS

FIG. 44-2

This circuit can be used to interface a 2-wire transmitter/sensor combination to an external device or measurement setup.

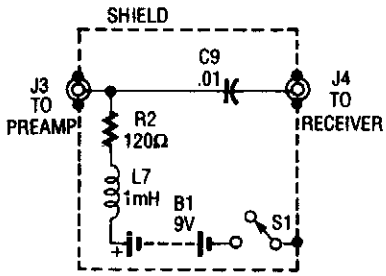
RELAY INTERFACE FOR AMATEUR RADIO TRANSCEIVERS



73 AMATEUR RADIO *FIG. 44-3*

The relay power in the linear is obtained from the –120-V bias supply, and the transmit keying output from the Kenwood is +12 V at 10 mA maximum. The key ingredient in the circuit is the pnp driver transistor, which must be capable of handling at least 150 V at about 250 mA.

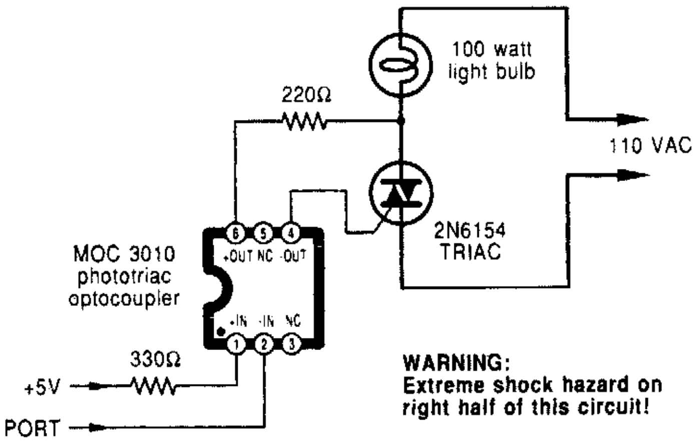
RECEIVER-INTERFACE CIRCUIT FOR PREAMPS



POPULAR ELECTRONICS *FIG. 44-4*

The purpose of the receiver/interface circuit is to pass RF to the receiver through capacitor C9, while adding dc power to the feedline through R2 and RF choke L7.

MICROCOMPUTER-TO-TRIAC INTERFACE



RADIO-ELECTRONICS

FIG. 44-5

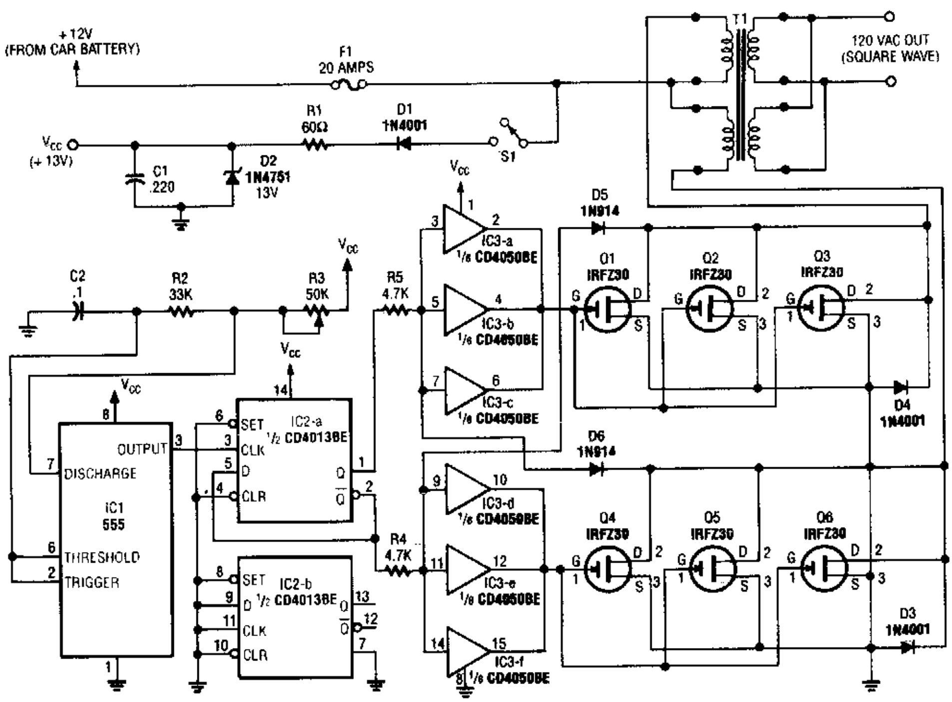
A microcomputer-to-triac interface uses a phototriac optoisolator to let safety-isolated logic signals directly control high-power loads. Depending on the input waveforms and the load, this circuit can be used in either an on/off switch or a proportional phase control. A low input powers the lamp.

45

Inverter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

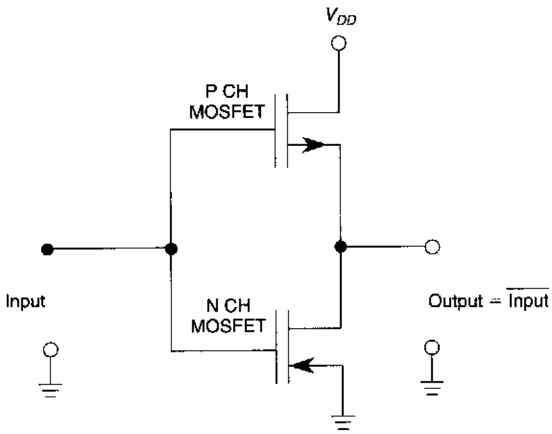
250-W INVERTER



ELECTRONICS NOW ***FIG. 45-1***

A 555 timer (IC1) generates a 120-Hz signal that is fed to a CD4013BE flip-flop (IC1-a), which divides the input frequency by two to generate a 60-Hz clocking frequency for the FET array (Q1 through Q6). Transformer T1 is a 12-/24-V center-tapped 60-Hz transformer of suitable size.

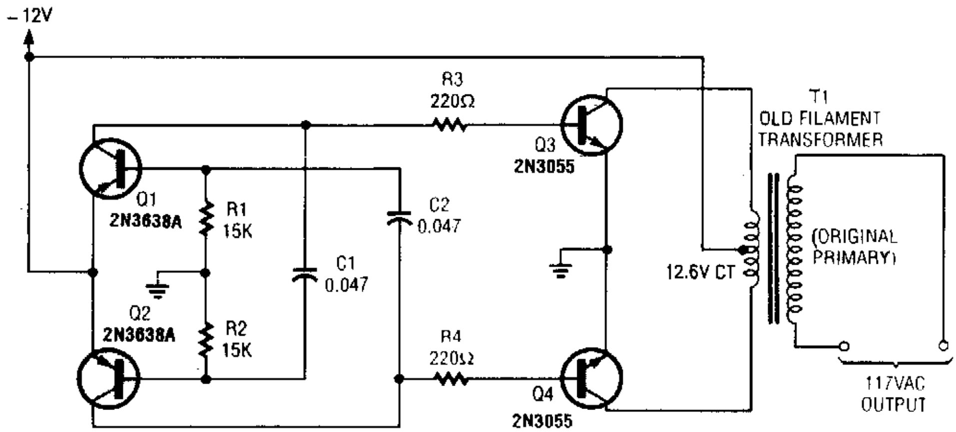
DIGITAL INVERTER



A CMOS digital inverter is formed by connecting two MOSFETS, as shown.

WILLIAM SHEETS ***FIG. 45-2***

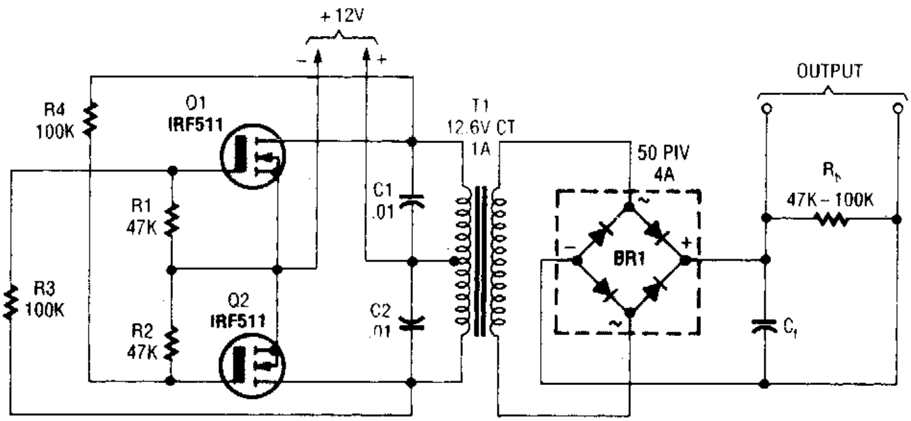
dc-to-ac INVERTER



POPULAR ELECTRONICS ***FIG. 45-3***

A multivibrator circuit drives a pair of 2N3055 power transistors. T1 is a 12.6-V CT filament transformer with a 120-V primary.

POWER MOSFET INVERTER



POPULAR ELECTRONICS ***FIG. 45-4***

T1 is a suitable transformer for the voltage desired, with a 12.6-V CT winding.

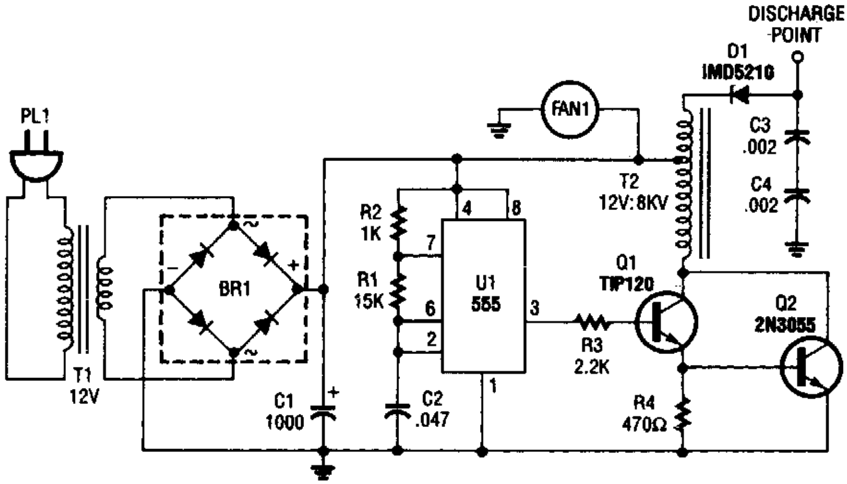
46

Ion Generator Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Negative Ion Generator

NEGATIVE ION GENERATOR



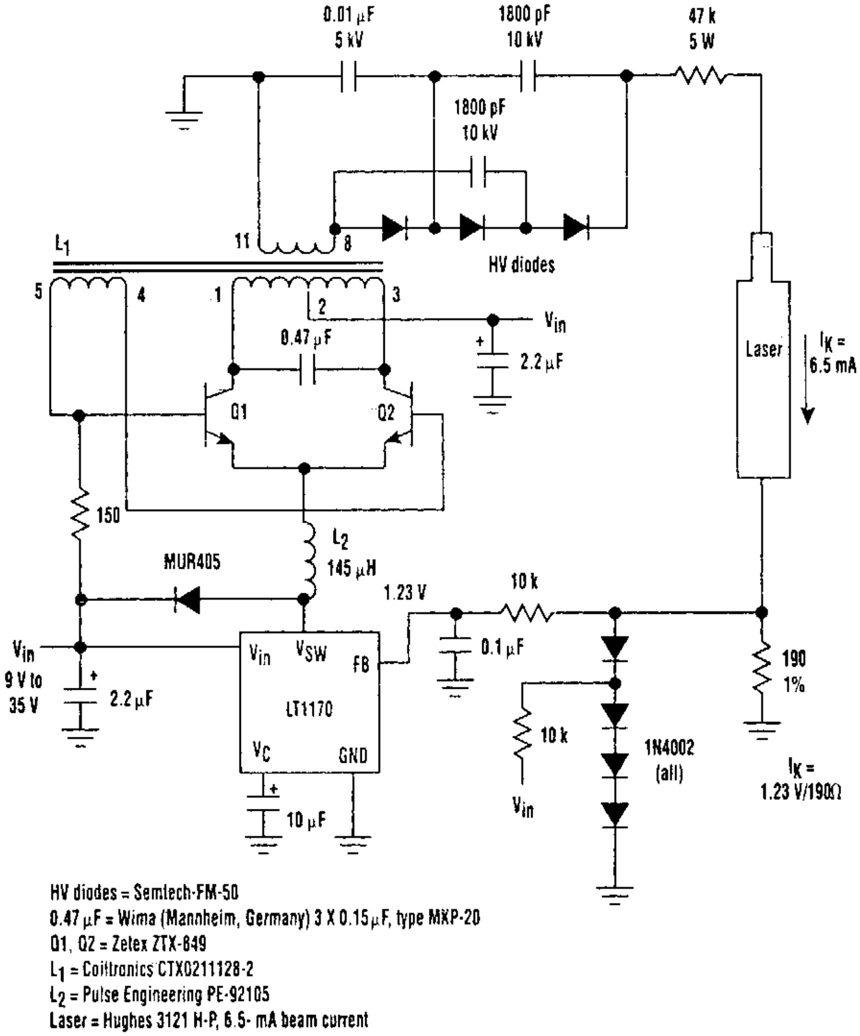
1993 ELECTRONICS HOBBYIST HANDBOOK *FIG. 46-1*

This oscillator-driver induces a high voltage in the windings of T2.

47

Laser Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

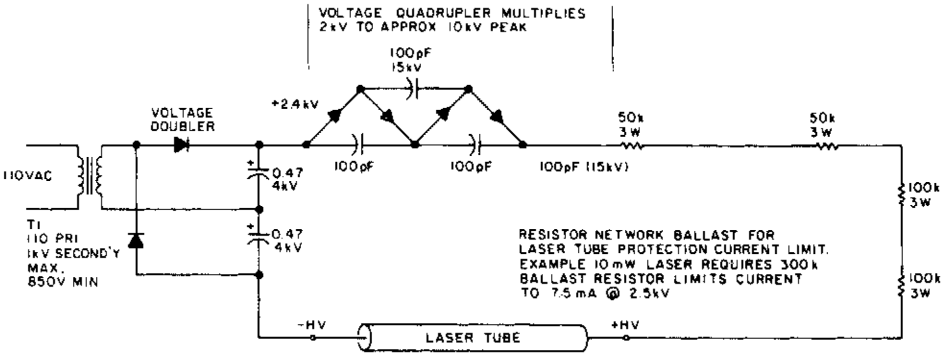


ELECTRONIC DESIGN

FIG. 47-1

Driving Helium-Neon Lasers can be simplified considerably using this power-supply configuration. When power is applied, the laser doesn't conduct and the voltage across the 190-Ω resistor is zero. However, a resonant circuit and a voltage tripler then produces over 10 kV to turn on the laser.

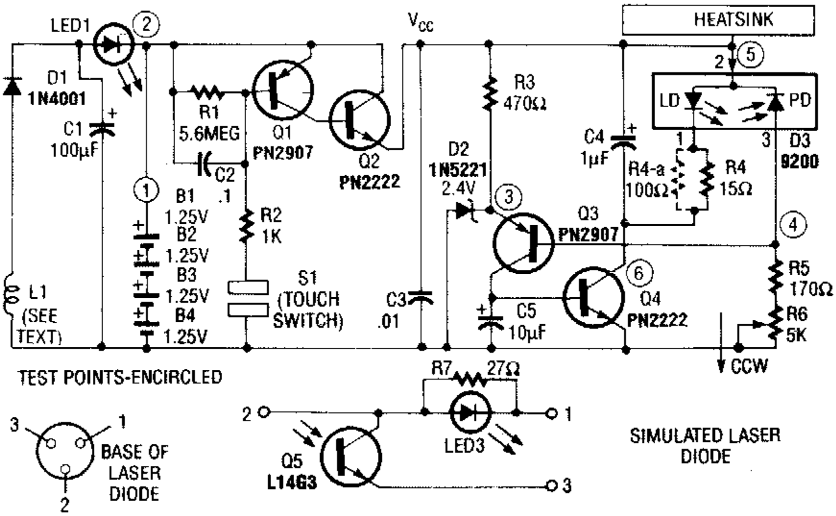
LASER POWER SUPPLY AND STARTING CIRCUIT



73 AMATEUR RADIO TODAY ***FIG. 47-2***

This circuit delivers 10 kV peak, then limits current to 7.5 mA @ 2 kV. The resistors shown provide ballasting. The starting circuit cannot maintain the 10 kV under load and appears as a series-pass circuit with little drop in voltage.

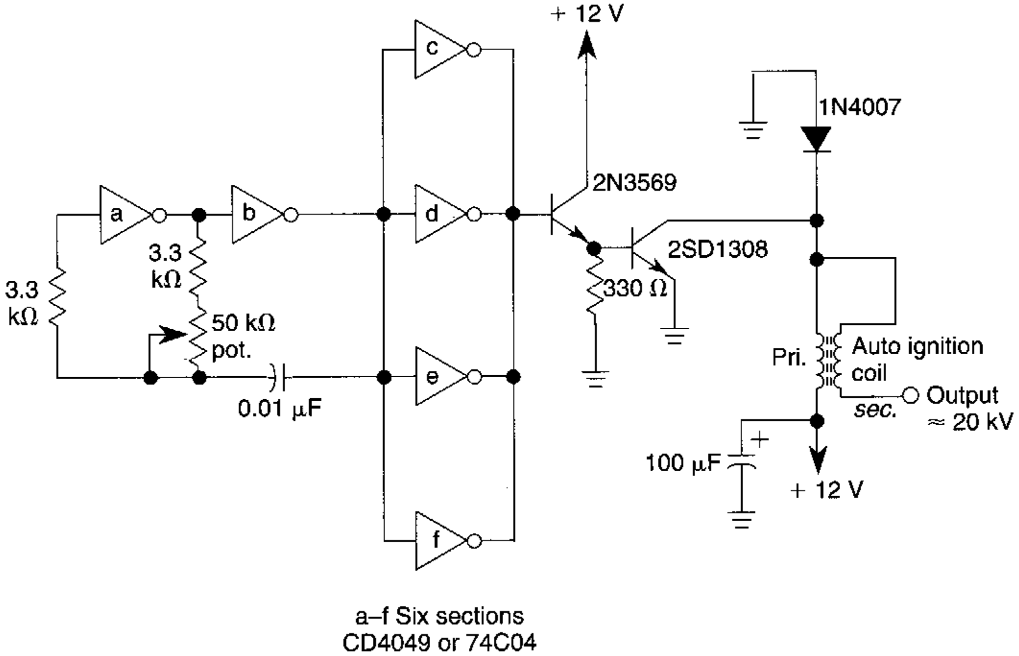
HANDHELD LASER



1992 R-E EXPERIMENTERS HANDBOOK ***FIG. 47-3***

A laser diode TOLD9200 (Toshiba) is used as a source of laser light. Q3, Q2, and S1 form a touch switch to control the laser. L1 is an RF pickup coil to pick up energy from an RF-type battery charger. It is 10 turns of #18 wire on a ½" diameter.

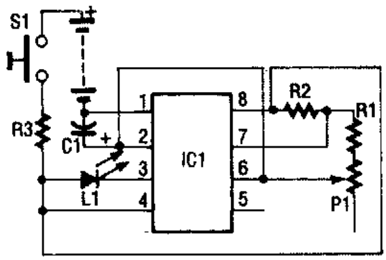
HIGH-VOLTAGE POWER SUPPLY



WILLIAM SHEETS ***FIG. 47-4***

The high-voltage power supply is a CMOS-based oscillator that pulses a high-voltage ignition transformer. The transformer output is around 20 kV.

FANTASTIC SIMULATED LASER

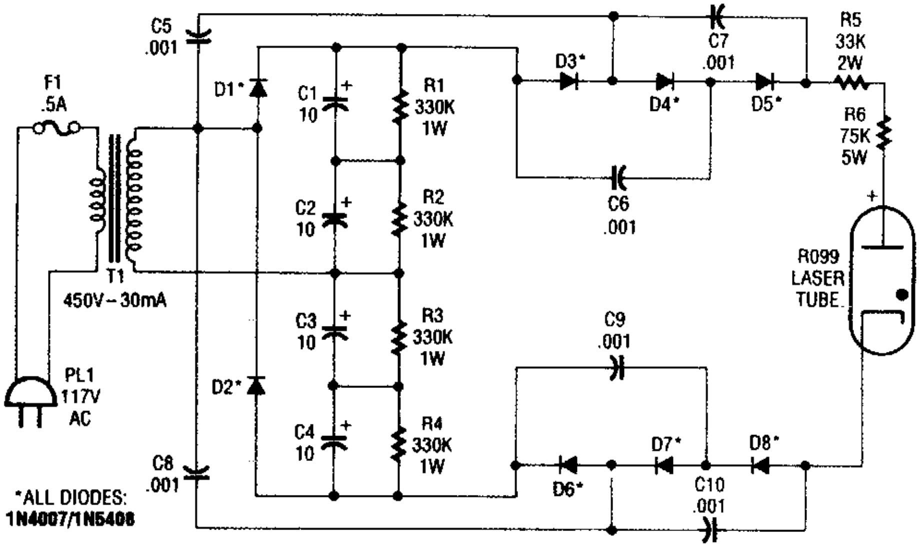


C1	= 1μF
IC1	= 555
L1	= Bright LED
P1	= 50 kΩ pot.
R1, R2	= 4.7 kΩ
R3	= NO push button

1991 PE HOBBYIST HANDBOOK ***FIG. 47-5***

The circuit uses a 555 timer IC to power an ultrabright LED. The output is a pulsing red light that can be projected using lenses. An ultrabright Stanley LED, capable of 300-millicandle output, is tied to pin 3 of the 555 timer IC. That IC has been configured as an astable multivibrator. The frequency of this multivibrator is controlled by R1, R2, C1, and P1. You can vary the frequency by adjusting P1, which changes the output from a slow blinking to a fast pulsating light. Resistor R3 is used to limit the current flowing into the circuit to a safe value, to prevent the LED and the IC from burning out. Switch S1 applies power to the circuit when its button is pressed.

LASER POWER SUPPLY



POPULAR ELECTRONICS *FIG. 47-6*

This supply generates an initial high voltage for ignition purposes. After ignition, the supply generates about 1300 to 1500 V. If a higher ignition voltage (than the 6000 V supplied) is necessary, more multiplier stages can be added to D5 and D8.

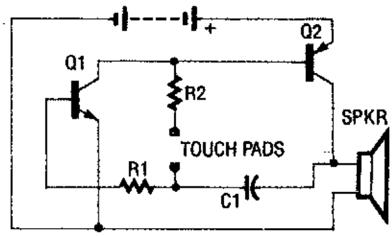
48

Lie Detector Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Simple Lie Detector

SIMPLE LIE DETECTOR



1991 PE HOBBYIST HANDBOOK

FIG. 48-1

The circuit uses a two-transistor direct-coupled oscillator that has a frequency determined by C1, R2, and the (skin) resistance across the touch pads. Since C1 and R2 are fixed values, only the skin resistance across the touch pads can vary the sound of the oscillator. To sustain oscillations, C1 feeds a portion of the output from Q2 back to the input of Q1 through resistor R1.

Transistor Q1 is an npn type and transistor Q2 is a pnp type. The output of Q2 is fed into a small speaker. The circuit relies on the fact that the human skin conducts electricity.

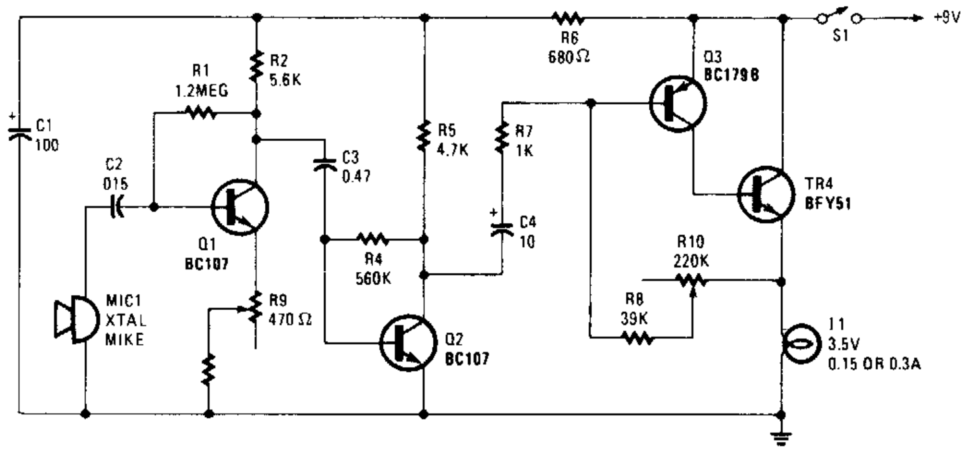
C1	0.01-µF Capacitor
Q1	2N3904 Transistor
Q2	2N3906 Transistor
R1	4.7 kΩ Resistor
R2	82 kΩ Resistor

49

Light Beam Communication Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

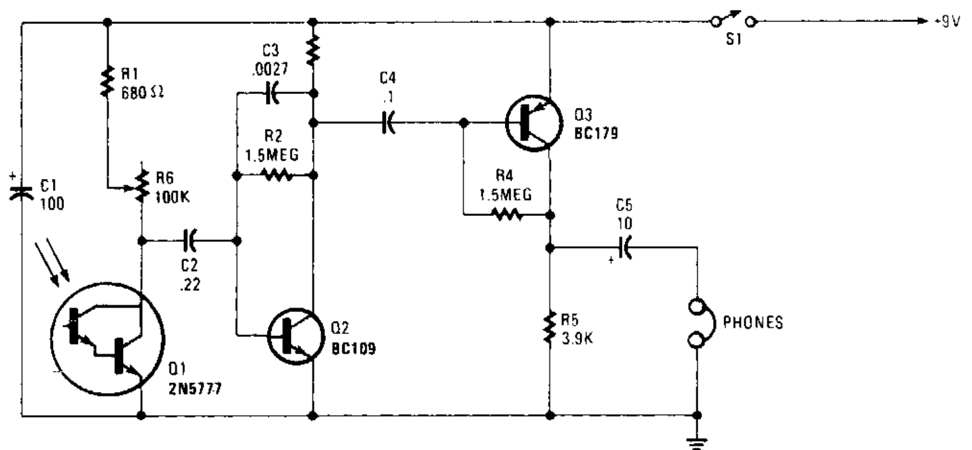
MODULATED LIGHT TRANSMITTER



POPULAR ELECTRONICS ***FIG. 49-1***

A light-bulb filament can be modulated with audio as a method of optical transmission. Amplifier Q1/Q2/Q3 drives emitter-follower TR4. Adjust R10 for the Q point (light bulb) giving best results. It should have a filament with low thermal inertia for best audio responses.

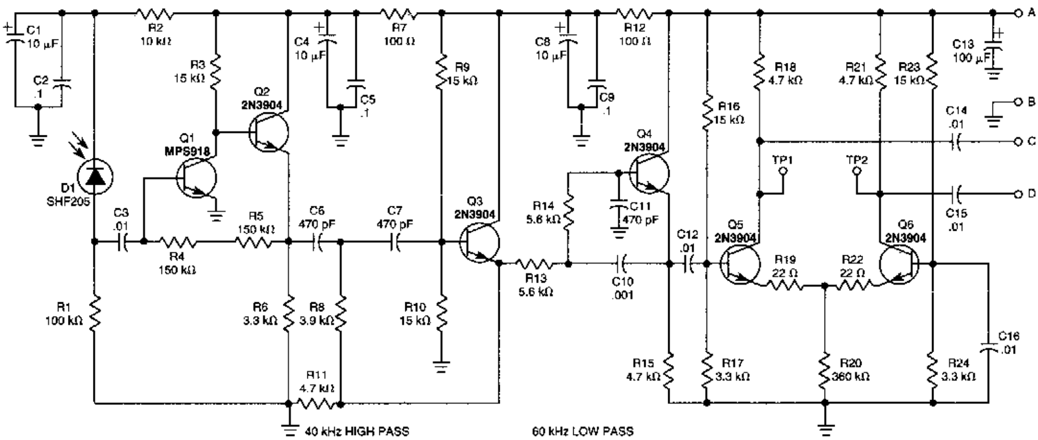
MODULATED LIGHT RECEIVER



POPULAR ELECTRONICS ***FIG. 49-2***

Using a phototransistor, this receiver will detect and demodulate a modulated light beam. R6 affects sensitivity.

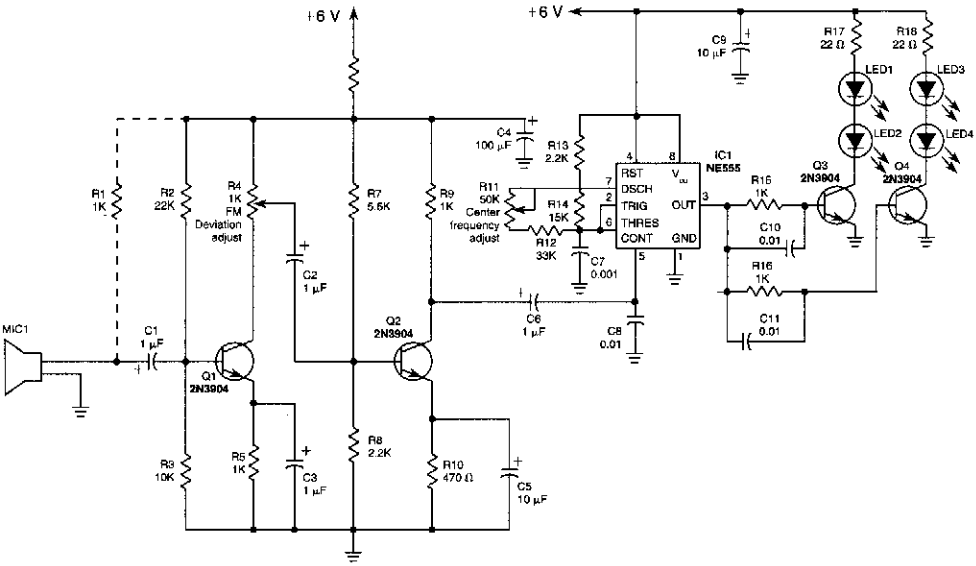
FM LIGHT BEAM RECEIVER



ELECTRONICS NOW ***FIG. 49-3***

This receiver will pick up IR or light beams that are frequency modulated on a 50-kHz carrier. Q2/Q1/Q3/Q4 from an active filter and amplifier and differential amp Q5/Q6 provide more gain.

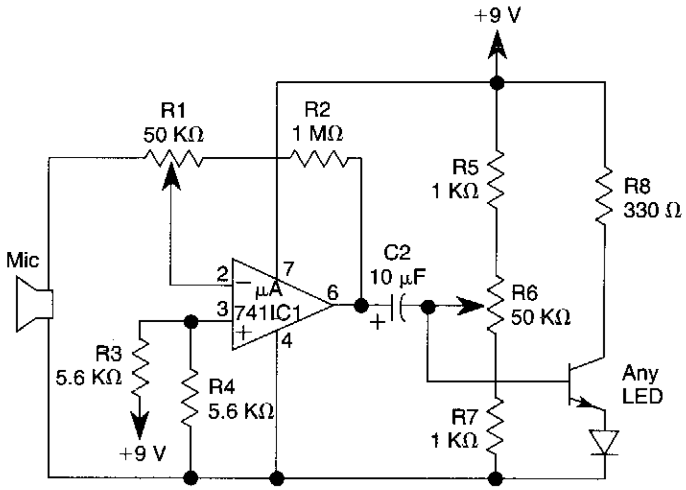
FM LIGHT-BEAM TRANSMITTER



ELECTRONICS NOW ***FIG. 49-4***

This transmitter uses two-stage amplifier Q1/Q2 to frequency modulate an NE555 (configured as a VCO) operating at about 50 kHz. The resultant FM-modulated pulse train is converted to light pulses via LED1 through LED4, driven by Q3 and Q4.

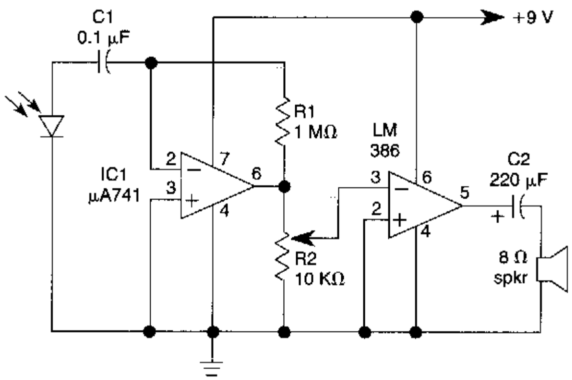
LIGHT-WAVE VOICE-COMMUNICATION TRANSMITTER



WILLIAM SHEETS **FIG. 49-5**

This transmitter uses a 741 op amp as a high-gain audio amplifier, which is driven by a microphone. The output of the 741 is coupled to Q1, which serves as the driver for a LED. Potentiometer R1 is the amplifier's gain control. Miniature trimmer resistor R6 permits adjustment of the base bias of Q1 for best transmitter performance. Gain control R1 can be eliminated if C1 and R2 are connected directly to pin 2 of the 741. For maximum sensitivity, increase the value of R_2 from 1 to 10 MΩ and use a crystal microphone with a large diaphragm.

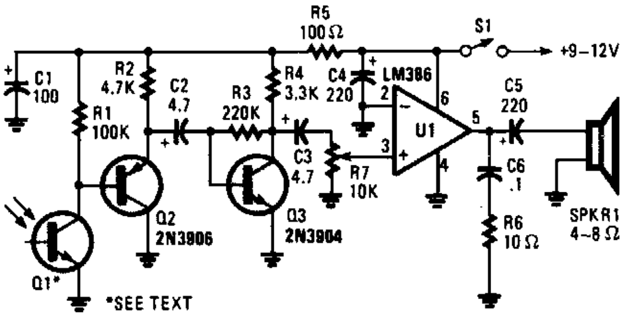
LIGHT-WAVE VOICE-COMMUNICATION RECEIVER



This light-wave receiver consists of a 741 operated as a preamplifier and an LM386 operated as a power amplifier. Potentiometer R2 is the gain control. Various kinds of detectors can be used as the front end of the receiver. Phototransistors are very sensitive, but they do not work well in the presence of too much ambient light. A 100-kΩ series resistor is required if you use a phototransistor. Solar cells, photodiodes, and LEDs of the same semiconductor as the transmitter all work well in this circuit.

WILLIAM SHEETS **FIG. 49-6**

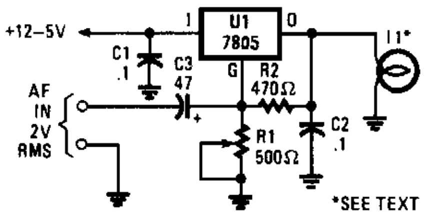
VISIBLE-LIGHT RECEIVER



POPULAR ELECTRONICS *FIG. 49-7*

This receiver for amplitude-modulated light signals uses phototransistor Q1 mounted in a parabolic reflector (to increase range). Any npn phototransistor should work. Emitter-follower Q2 drives amplifier Q3. The output from Q3 feeds volume control R7 and audio amplifier U1. A 9- to 12-V supply is recommended for the receiver.

VISIBLE-LIGHT AUDIO TRANSMITTER



In the visible-light transmitter, a 7805 voltage regulator is connected in a variable-voltage configuration, and an audio signal is fed to the common input, to modulate the output voltage. The modulated output voltage is used to transmit intelligence via an incandescent lamp.

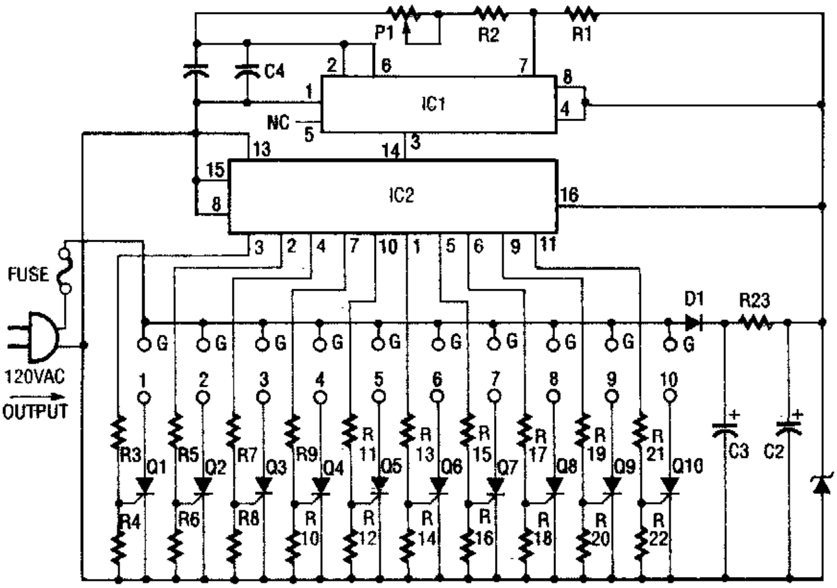
POPULAR ELECTRONICS *FIG. 49-8*

50

Light Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

LIGHT SEQUENCER



1991 PE HOBBYIST HANDBOOK ***FIG. 50-1***

The light sequencer uses two ICs and 10 SCRs to create an ac sequencer. The first IC, a 555 timer, is used to provide clock pulses for IC2. The IC is configured as an astable multivibrator, and its output is on pin 3.

Capacitors C1 and C4, along with resistor R2 and potentiometer P1, control the frequency of the pulses. IC2 is a 4017 Johnson counter, which shifts a high-signal level to each one of its 10 output pins in sequence. Each output pin is resistively coupled to the gate lead on an SCR. When the respective output pin on the 4017 is high and the positive half of the ac cycle is on the anode lead of the SCR, it turns on. The lamp that is connected to its anode lights.

Power is brought into the PC board by the line cord, then the circuit is fuse-protected. Diode LD1 changes the ac to pulsating, which is smoothed by C2 and C3. R23 limits the current, and zener diode D2 limits the dc voltage to 6 Vdc.

CI, C4	0.1-µF Capacitor	R2, R4, R6,	
C2	100-µF Capacitor	R8, R10, R12,	
C3	47-µF, 350-V Electrolytic Capacitor	R14, R16, R18	
D1	1N4007 Diode	R20,R22	100-kΩ Resistor
D2	6-V Zener (M747814)	R3, R5, R7	
IC1	555 Timer IC	R9, R11, R13	
IC2	4017 CMOS IC	R15, R17, R19	
P1	500-kΩ Potentiometer	R21	2.2-kΩ Resistor
Q1–Q10	106 SCR	R23	15-kΩ 7-W Resistor
R1	560-Ω Resistor		

HOLIDAY LIGHT SEQUENCER

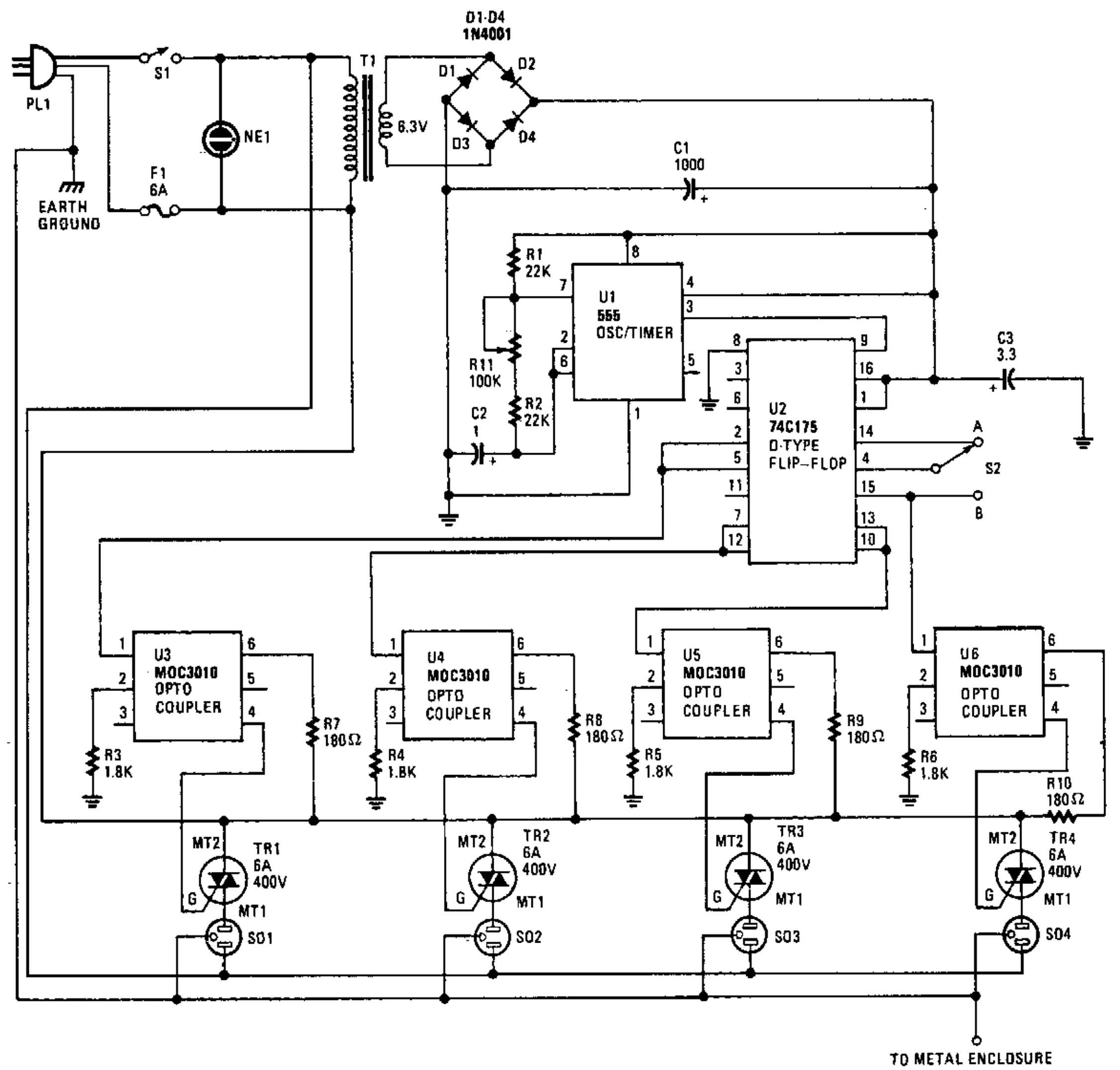


FIG. 50-2

Integrated circuit U1 (a 555 oscillator/timer) is wired as a conventional pulse generator. The frequency of the pulse generator is controlled by potentiometer R11. Resistor R2 puts a reasonable limit on the highest speed attainable.

The output of the pulse generator is fed to the common clock input of U2, a 74C175 quad D-type flip-flop. Each flip-flop is configured so that its Q output is coupled to the D input of the subsequent flip-flop.

Information on the D input of each flip-flop is transferred to the Q (and Q) outputs on the leading edge of each clock pulse. Switch S2 allows you to invert the information on the D input of the first flip-flop at any time during the cycle. This allows you to create a number of different sequences, which are determined by the state of the CQ output at the time of the switching.

Some of the possible sequences are:

- 1 through 4 on, 1 through 4 off;
- 1 of 4 on sequence;
- 1 of 4 off sequence;
- 2 of 4 on sequence;
- 1 and 3 on to 2 and 4 off;
- and other instances when the sequence of events is difficult to determine.

However, if S2 is switched to position B while all outputs are high or all are low (which seldom occurs), the sequence stops and the outputs remain either all on or all off. If that happens, you only need to switch back to position A for at least one pulse duration, then back to position B again.

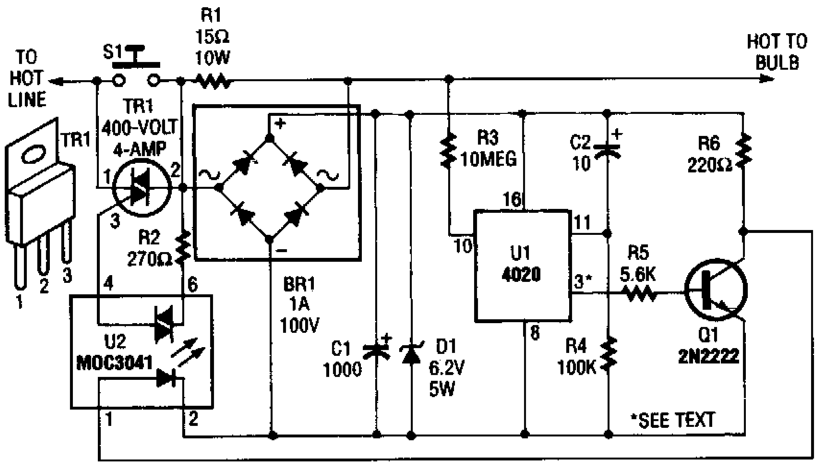
Likewise, S2 should be in position A (pin 4 connected to pin 14) each time the power is turned on. This is because the data on pin 4 must be a logic 1 in order to start a sequence; otherwise all outputs remain at logic 0, regardless of the clock pulses.

Each output of the sequencing circuit is connected to an MOC3010 optoisolator/coupler (U3 through U6), which contains an infrared-emitting diode with an infrared-sensitive diac (triac driver or trigger) in close proximity. The diac triggers the triac, which carries the 117-volts ac.

Each time that the infrared-emitting diode receives a logic 1, it turns on and causes the diac to conduct. With the optoisolator/coupler's internal diac conducting, the triac turns on, and power is supplied to whatever load is plugged into the corresponding ac socket. So, the sequencing circuit and the 117-V ac outputs are "optically coupled" and are effectively isolated from each other.

Power for the sequencing circuit is provided by a 6.3-V miniature transformer. The output of the transformer is rectified by a four-diode bridge circuit, the output of which is filtered by C1 (1000-μF electrolytic capacitor). Capacitor C3 is added at the supply pin of U2 to suppress transients.

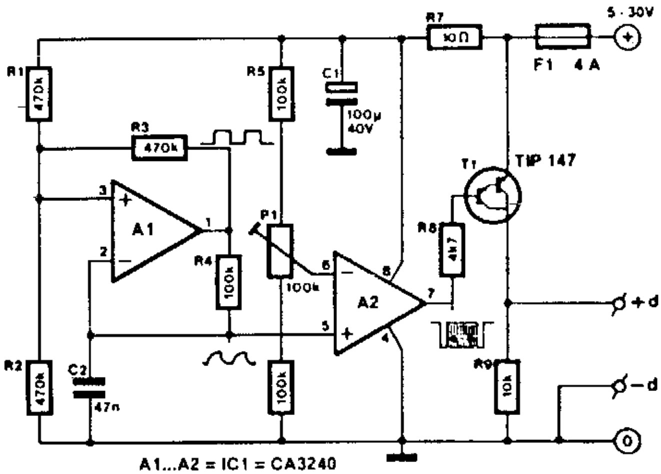
AUTOMATIC PORCH-LIGHT CONTROL



1993 ELECTRONICS HOBBYISTS HANDBOOK ***FIG. 50-3***

The automatic porch-light control circuit holds a triac on until a 4020 divider counts a number of 60-Hz powerline pulses. The circuit turns off a light after a predetermined time by using pins other than pin 3 of U1. Various times can be set. Consult the 4020 data sheet for information.

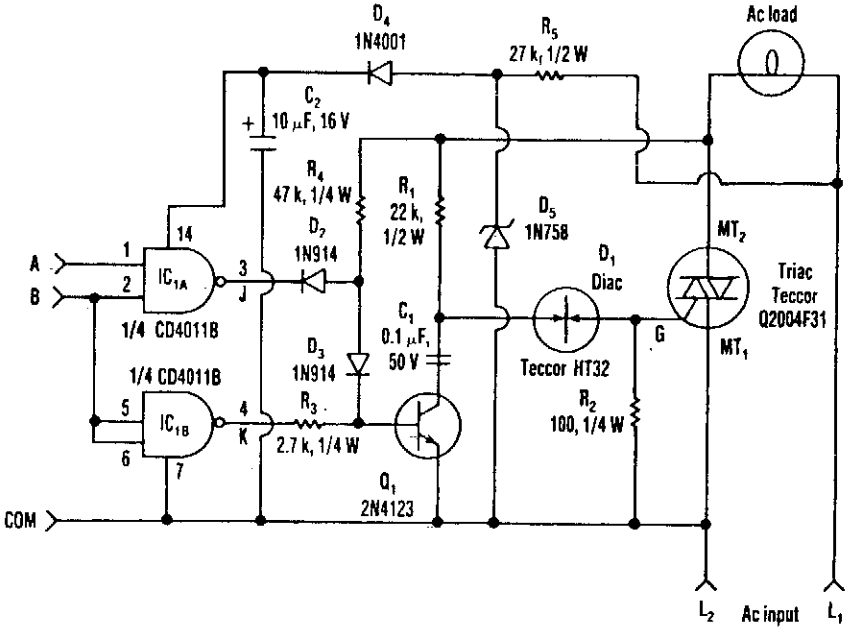
DIMMER FOR LOW VOLTAGE LOADS



303 CIRCUITS ***FIG. 50-4***

This circuit controls a low voltage dc supply by pulse width modulation. The switching rate is 200 Hz. Input supply voltage should be +5 to +30 V. Up to 5 A can be controlled.

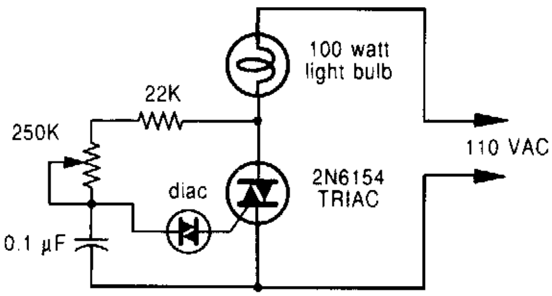
THREE-POWER-LEVEL TRIAC CONTROLLER



ELECTRONIC DESIGN ***FIG. 50-5***

Three power levels are supplied by the two logic inputs of this enhanced circuit. R5, D4, D5, and C2 form a power supply for the logic IC. They can be omitted if another source of low voltage is available.

PHASE-CONTROLLED DIMMER

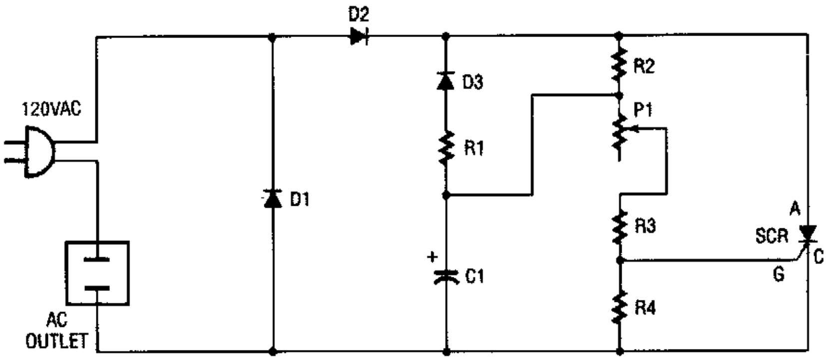


WARNING: Extreme shock hazard!

RADIO-ELECTRONICS ***FIG. 50-6***

A phase-controlled dimmer delays the triac turn-on to a selected point in each successive ac half cycle. Use this circuit only for incandescent lamps, heaters, soldering irons, or "universal" motors that have brushes.

120-ac SHIMMERING LIGHT



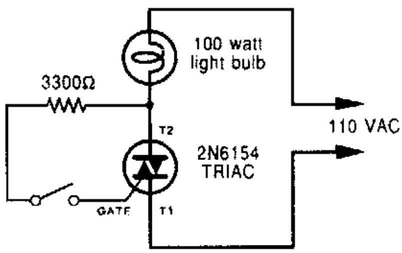
1991 PE HOBBYIST HANDBOOK

FIG. 50-7

You can turn any ordinary household bulb into one that shimmers or blinks. This circuit works on any incandescent light up to 200 W, and runs on standard 120 Vac. The circuit uses an SCR to cause an ordinary lamp to shimmer. Note that one side of the lamp is connected directly to 120 Vac, and the other side of the lamp goes to the cathode of the SCR. As ac voltage is brought into the circuit through the line cord, it is full-wave rectified by diodes D1 and D2. That changes the ac to dc, and a portion of that dc voltage is applied to capacitor C1 through R2. Diode D3 blocks the (+) dc voltage so that only the voltage from the path of R1 and D3 is clear. That forms an oscillator, which has a frequency determined by the setting of potentiometer P1 (because the other components have fixed values).

Remember to use **extreme caution** when using a device that connects to the ac line. **Never** use it outside or near water and always mount the entire kit inside a wooden or plastic (insulated) box to prevent any contact with the ac voltage.

SIMPLE TRIAC CIRCUIT



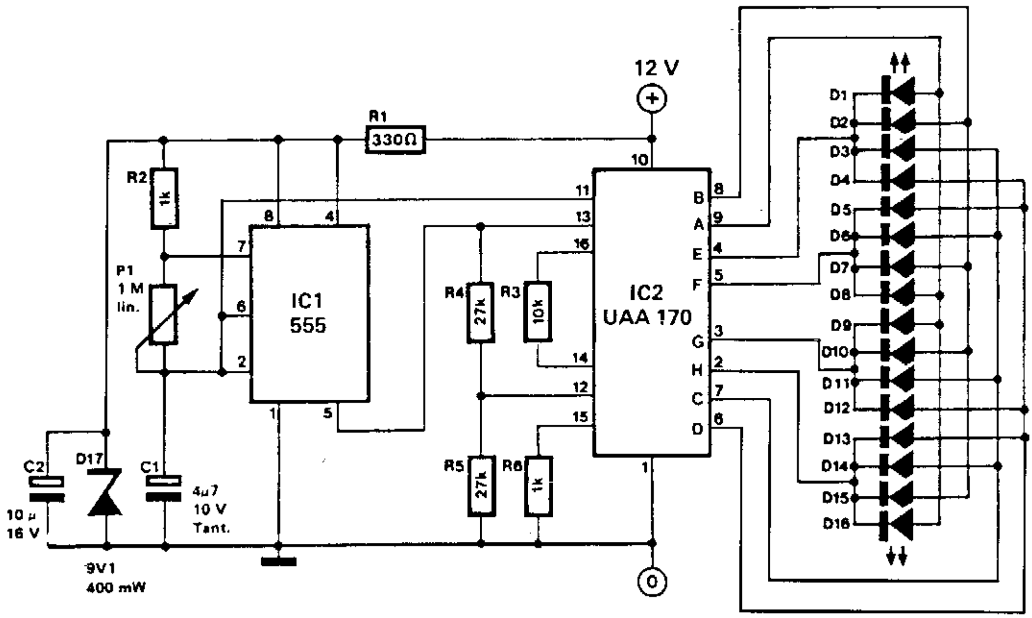
WARNING: Extreme shock hazard!

A triac can be used as a line-operated ac power switch that can directly control lamps, heaters, or motors. A brief and small current pulse into the gate turns the triac on; it remains on until the main current reverses.

RADIO-ELECTRONICS

FIG. 50-8

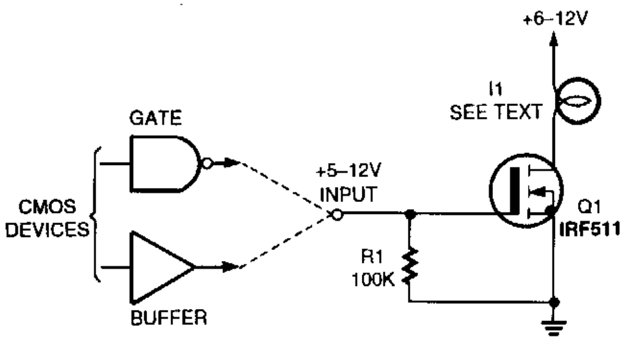
RUNNING LIGHT SEQUENCE



303 CIRCUITS ***FIG. 50-9***

This running light sequencer drives 16 LEDs and runs from a 12-V supply. C1 can be varied to alter the rate of operation.

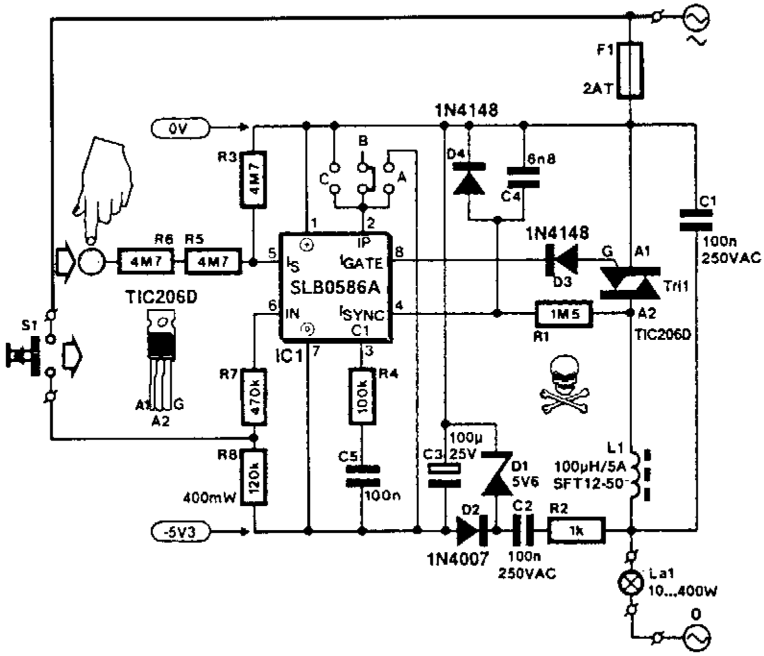
MOS LAMP DRIVER



POPULAR ELECTRONICS ***FIG. 50-10***

The circuit shows a way of using a MOSFET as a load driver. I1 can be a lamp, or any other load, that does not exceed the current rating of Q1.

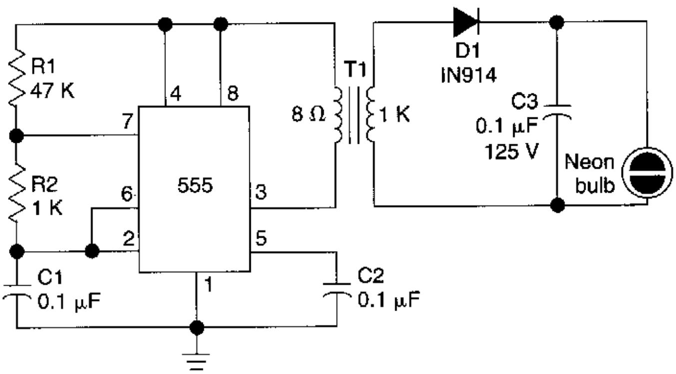
CMOS TOUCH DIMMER



ELEKTOR ELECTRONICS *FIG. 50-11*

A Seimens SLB0586A IC allows the construction of a simple touch-controlled dimmer circuit. The circuit controls a triac ac switch, which allows control of loads from 10 to 400 W.

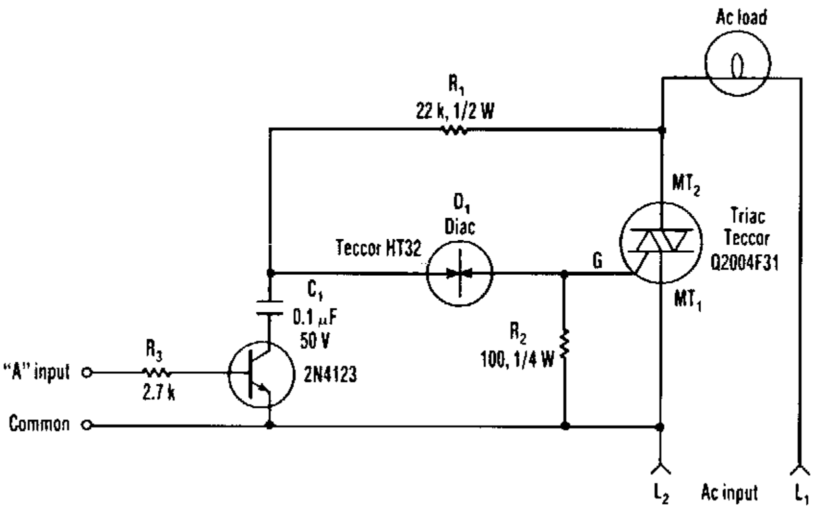
NEON LAMP DRIVER FOR 9-V SUPPLIES



RADIO-ELECTRONICS *FIG. 50-12*

This circuit is for driving a neon lamp from a 9-V supply. The 555 generates an ac signal (stepped up by T1), and lights the neon bulb. T1 is any small audio output transformer.

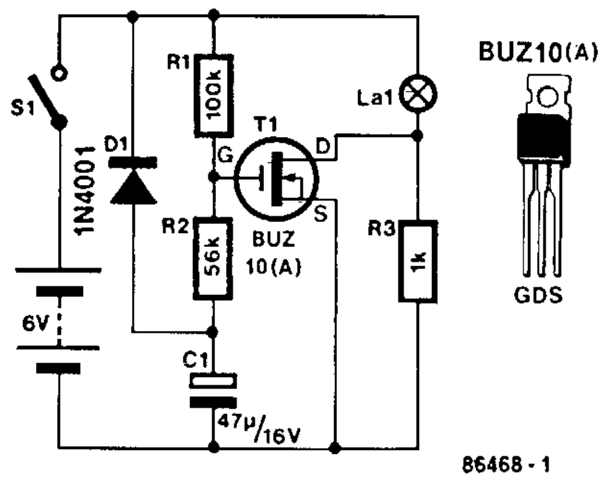
SENSITIVE TRIAC CONTROLLER



ELECTRONIC DESIGN *FIG. 50-13*

The single transistor connected between the capacitor and the common side of the ac line allows a logic-level signal to control this triac power circuit. Resistor R2 prevents false triggering of the triac by the trickle current through the diac.

HALOGEN LAMP PROTECTOR



303 CIRCUITS *FIG. 50-14*

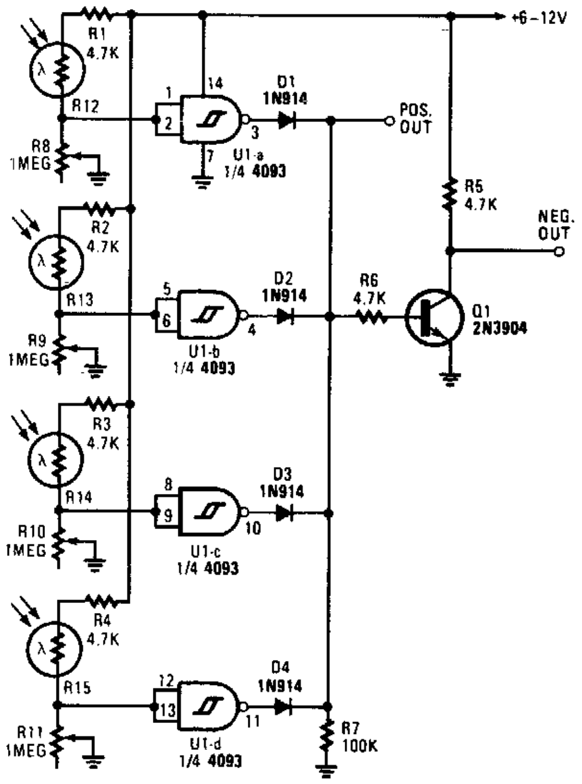
This circuit produces a soft turn-on for halogen lamp filaments upon powering up. MOSFET used is a BUZ10, which has 0.2 Ω R_{DS} on. R1, R2, and C1 set the turn-on rate and D1 discharges C1 at turn-off.

51

Light-Controlled Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

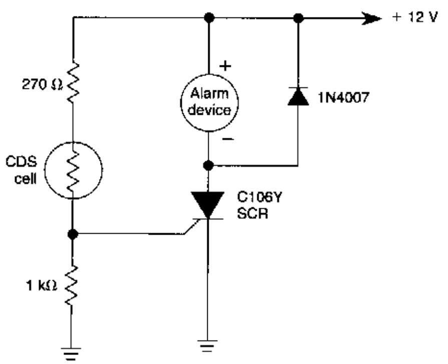
LIGHT-DEPENDENT SENSOR FOR MULTIPLE INPUTS



This light-dependent sensor uses LDRs to detect the presence or absence of light. As long as the light source striking the LDRs remains constant, the alarm does not sound. But when the light is interrupted, the alarm is triggered.

POPULAR ELECTRONICS *FIG. 51-1*

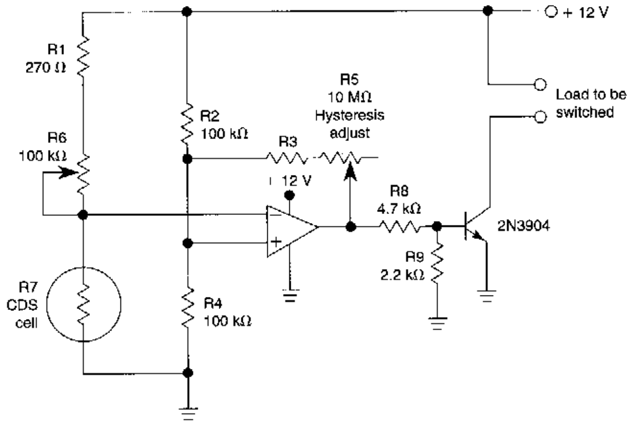
SIMPLE LIGHT-ACTIVATED ALARM



A cadmium-sulfide photocell conducts when a light beam strikes it. This triggers the SCR and activates the alarm device.

WILLIAM SHEETS *FIG. 51-2*

PRECISION DARK-ACTIVATED SWITCH WITH HYSTERESIS

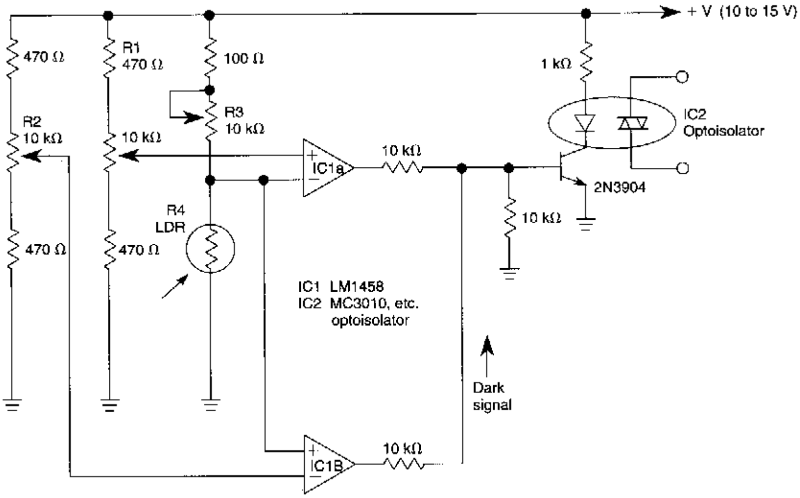


WILLIAM SHEETS

FIG. 51-3

A CdS cell is one leg of a bridge circuit. Potentiometer R6 in another leg sets the trip point. Potentiometer R5 provides hysteresis adjustment to prevent "chattering" or hunting of the relay. The light level has to increase noticeably before the 2N3904 turns off and the circuit deactivates.

COMBINED LIGHT-/DARK-ACTIVATED SWITCH



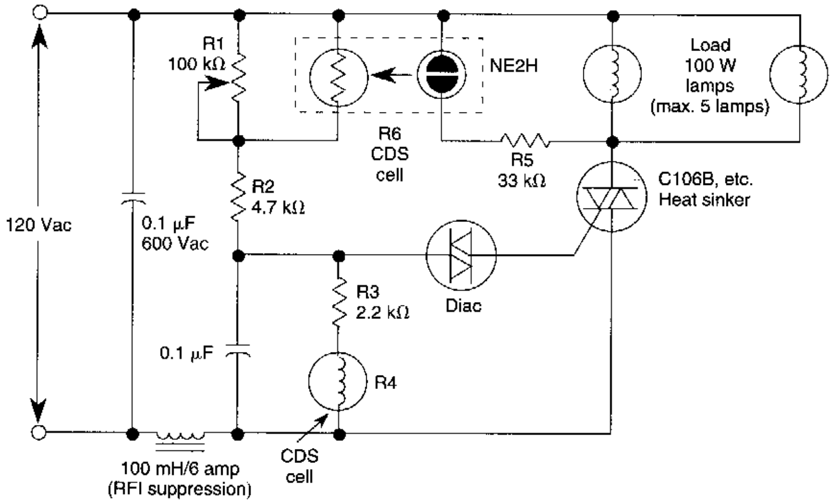
Set R4 so 1/2 of V_{CC} appears across R3.
Set R2 for dark trip point.
Set R1 for light trip point.

WILLIAM SHEETS

FIG. 51-4

Two op amps used in a bridge circuit configuration detect high and low light levels. Potentiometer R2 sets the dark level and R1 controls the light level. R3 is set so that about ½ the supply voltage appears across R4 at the desired light level. R1 and R2 set the trip point of the optoisolator IC2 at darker or lighter ambient levels, as required.

OUTDOOR LIGHT CONTROLLER



WILLIAM SHEETS

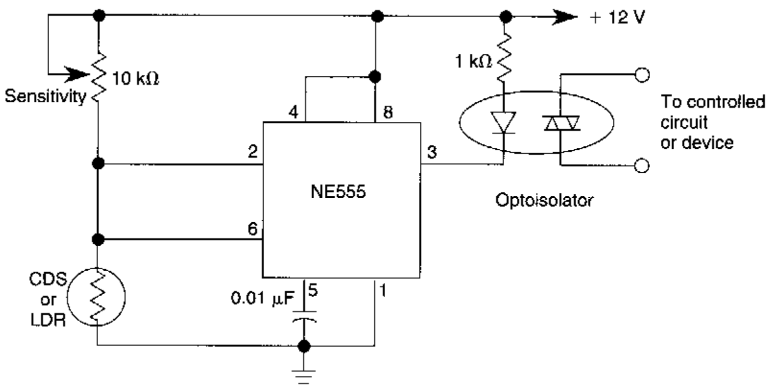
FIG. 51-5

A neon bulb and a CdS photocell enclosed in a light-tight enclosure form an optocoupler. A diac/triac combination is used to provide the snap-switch effect. A second CdS photocell acts as the main sensor.

As darkness approaches, the resistance of R4 begins to increase. At a threshold level, the diac triggers the triac and causes the neon bulb to light. This reduces the resistance of R6, causing the diac to trigger the triac, which lights the neon bulb and provides power to the load.

As morning light comes up, the process is reversed. The neon bulb goes out and the SCR turns off.

DARK-ACTIVATED RELAY WITH HYSTERESIS

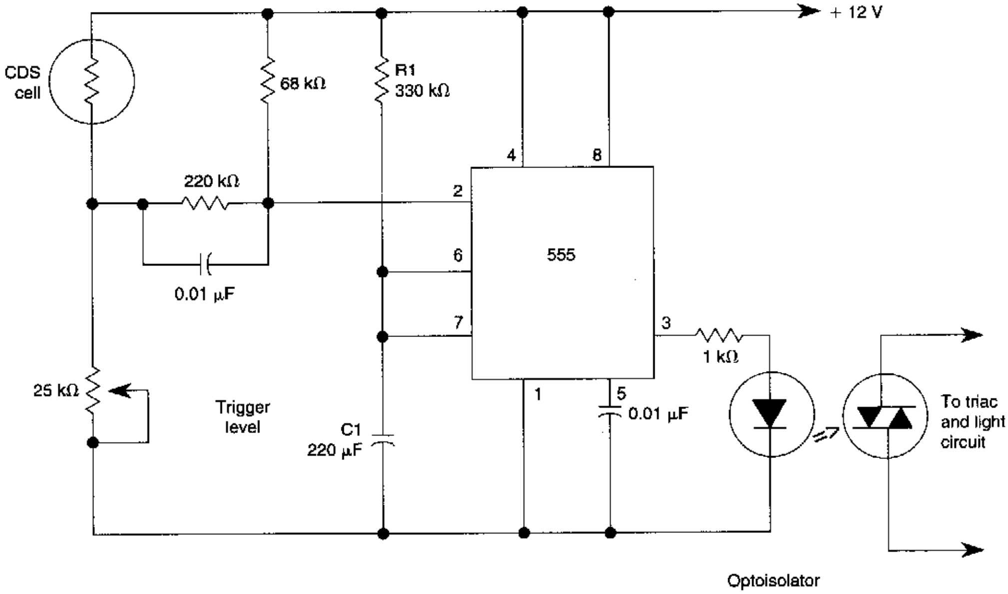


WILLIAM SHEETS

FIG. 51-6

The hysteresis of a 555 IC can be used to advantage for sensing a drop in light. An LDR or CDS cell with about 2 to 8 k resistance at desired light level should be used.

PORCH LIGHT CONTROL

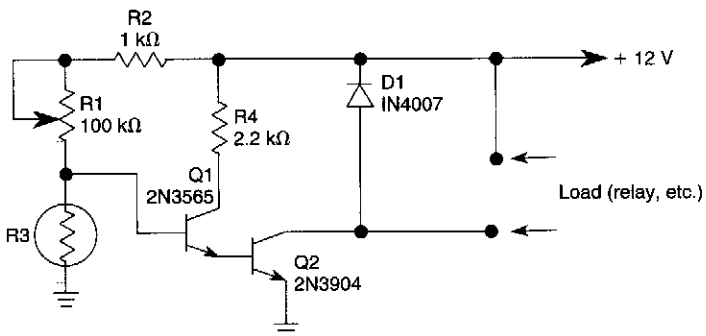


WILLIAM SHEETS

FIG. 51-7

This circuit can control the on/off cycle of a light via a CDS photocell, and turn it off after a preset period. The light can only be turned on when CDS cell is in darkness, and it stays on for a time determined by the 555 circuit. On time depends on R1 and C1 and is about 80 seconds with the values shown.

DARK-ACTIVATED SWITCH

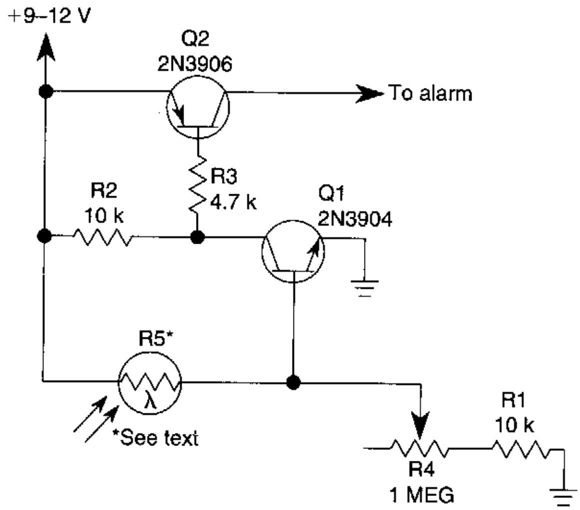


WILLIAM SHEETS

FIG. 51-8

In this circuit, lowering of the light level on the CDS cell turns on Q1 and Q2 which switches on the load which could be a relay, light, etc.

PHOTOELECTRIC SENSOR



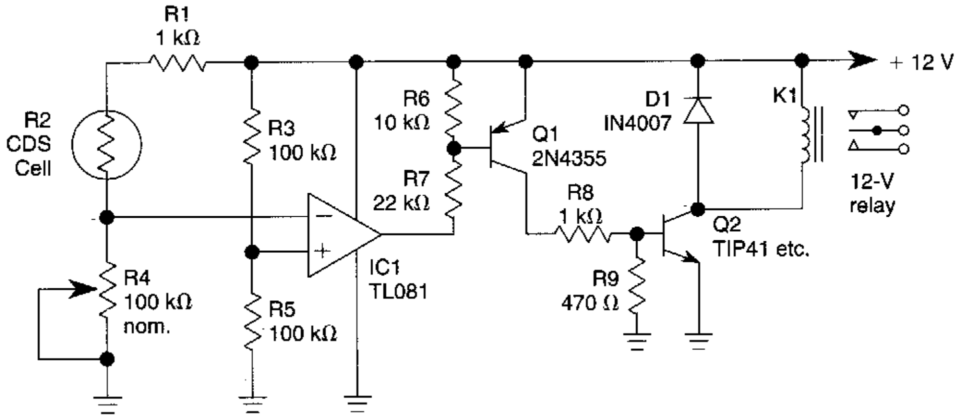
POPULAR ELECTRONICS ***FIG. 51-9***

The circuit can be used as a sensor that can trigger an alarm without direct contact being made by the intruder. In this circuit, a visible or invisible light source radiates on the sensor, keeping the detection loop in what could essentially be called a normally closed condition.

As long as the light source striking R5 remains uninterrupted, the switch remains closed. But if an intruder passes between the light source and the sensor, the circuit goes from closed to open, and triggers the alarm.

A light-dependent resistor (LDR), whose resistance varies inversely in with the amount of light hitting its sensitive surface, is used. A bright light aimed at R5 causes its internal resistance to drop as low as a few hundred ohms; in total darkness, the unit's resistance can rise to several megohms. The light-dependent resistor (R5) is connected between the +*V* supply and the base of Q1. As long as R5 detects light, it supplies ample base current to cause Q1's collector to saturate to near ground level. That also pulls the base of Q2 (a 2N3906 general-purpose pnp transistor) to near ground level, turning it on and clamping its collector to the +*V* rail.

PRECISION LIGHT-SENSITIVE RELAY SWITCH

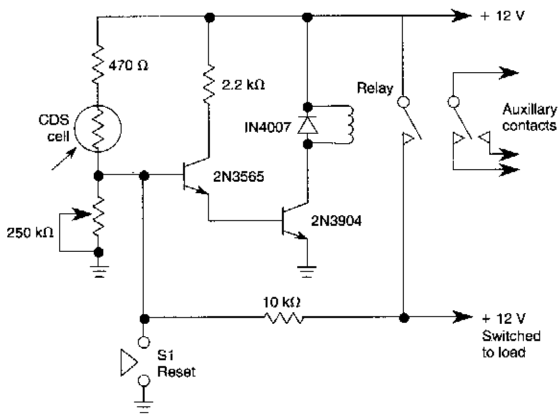


WILLIAM SHEETS

FIG. 51-10

A CDS cell in a bridge circuit with an op amp provides a simple means of operating a relay at a predetermined light level. Potentiometer R4 sets the sensitivity.

SELF-LATCHING LIGHT-ACTIVATED SWITCH

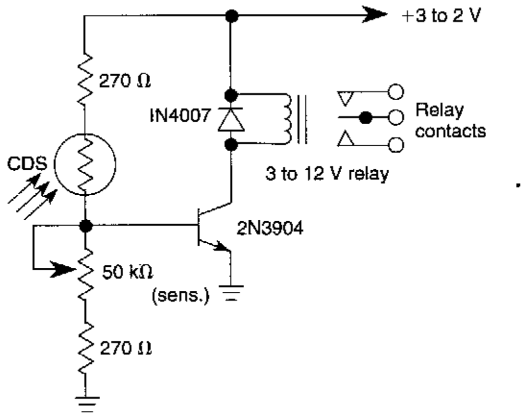


WILLIAM SHEETS

FIG. 51-11

When light strikes the CDS cell it turns on the transistors which activates the relay which latches. Depressing S1 grounds the base of the 2N3565 and the relay resets. The 250 k potentiometer adjusts the sensitivity of the circuit.

SIMPLE NONLATCHING PHOTOCELL SWITCH

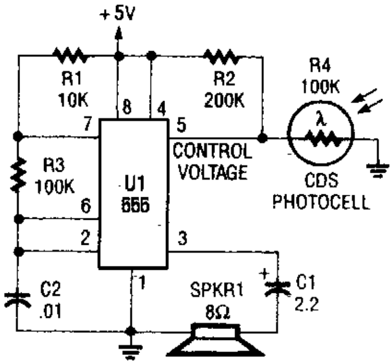


WILLIAM SHEETS

FIG. 51-12

A CDS photocell is used to drive the relay. The circuit operates from a +12 V supply.

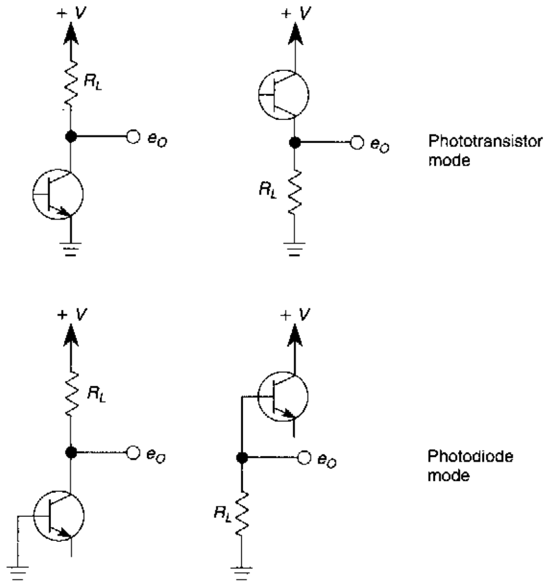
LIGHT-CONTROLLED OSCILLATOR



POPULAR ELECTRONICS ***FIG. 51-13***

This circuit can be used as a light detector and possibly as an aid for the visually handicapped. The frequency of the oscillator is determined by the amount of illumination striking LDR4.

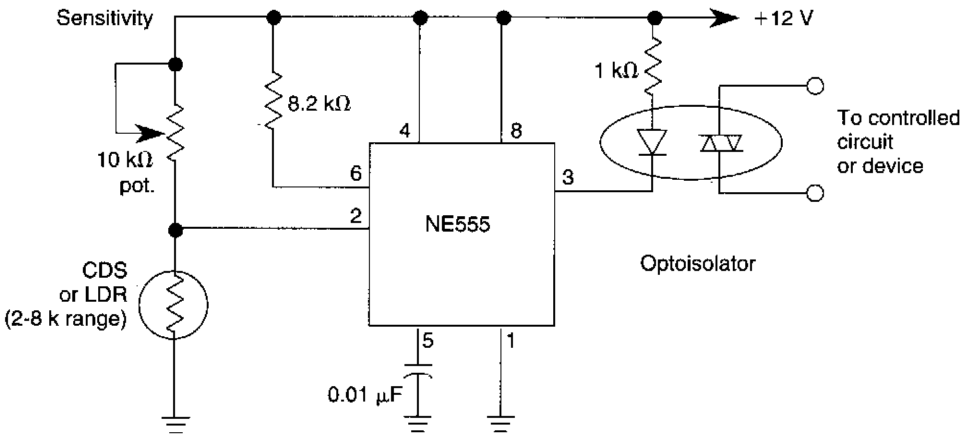
PHOTOTRANSISTOR CIRCUITS



WILLIAM SHEETS ***FIG. 51-14***

Here are four ways to connect a phototransistor for general use in phototransistor circuits.

DARK-ACTIVATED RELAY



WILLIAM SHEETS ***FIG. 51-15***

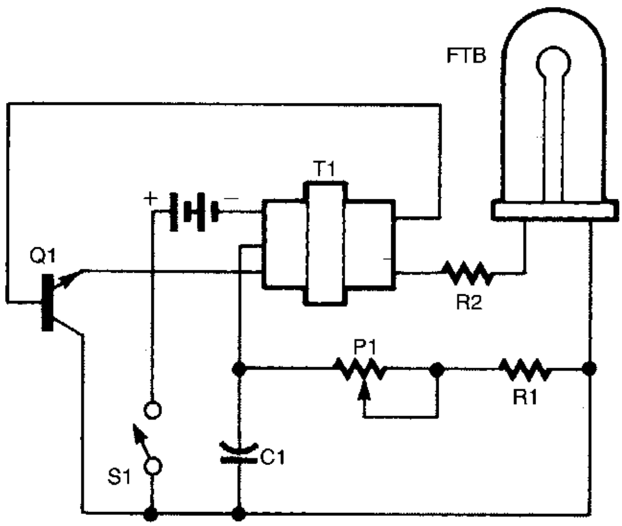
Configuring a 555 IC as shown yields a dark-activated relay with low hysteresis. CDS or LDR should be in the 2 k to 8 k range at desired light level.

52

Light Sources

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

BATTERY-OPERATED BLACK LIGHT

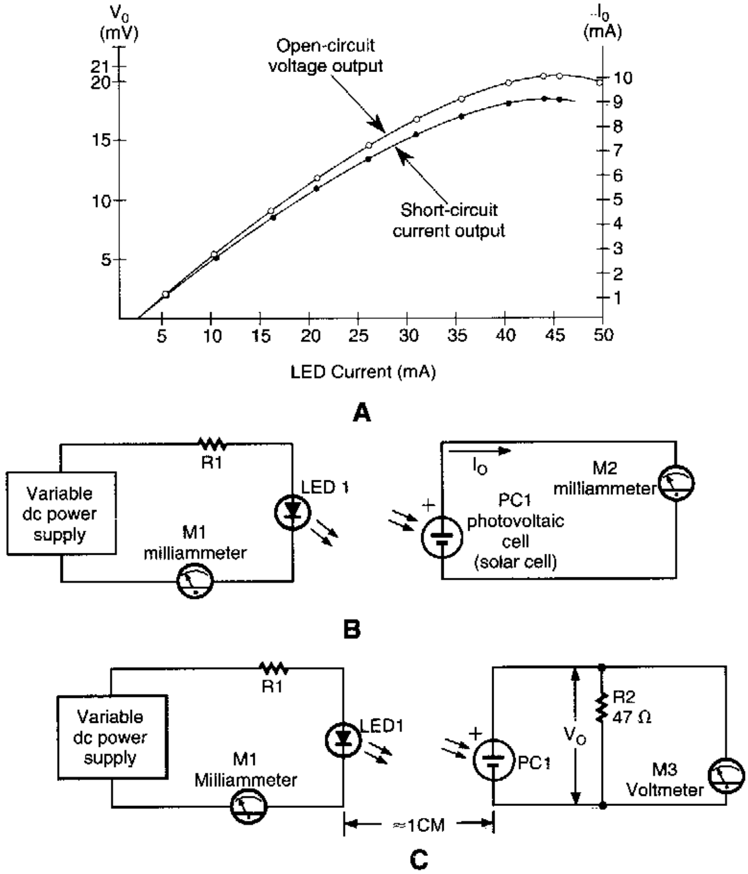


C1 – 1 μF Electrolytic Capacitor
FTB – FTB8T6 Black-light Tube
P1 – 2.5-k Trimmer Resistor (252C)
Q1 – C1740 SW Transistor
R1 – 330 Ω Resistor
R2 – 1-k Resistor
S1 – Slide Switch
T1 – Inverter Transformer

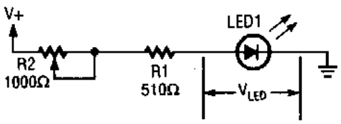
1989 PE HOBBYIST HANDBOOK ***FIG. 52-1***

The battery-operated black light uses a "U"-shaped, unfiltered, black-light tube, which requires approximately 250 Vac to operate. To create the 250-Vac 6-V battery, the circuit uses a one-transistor blocking oscillator that drives a ferrite inverter transformer. A blocking oscillator turns itself off after one or more cycles. In this circuit, it consists of C1, P1, Q1, R1, and T1. The oscillations are sustained because the base of Q1 is connected to one of the windings on T1.

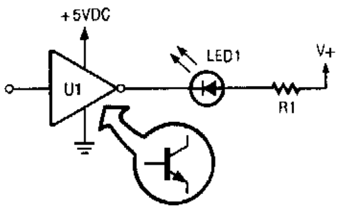
Transformer T1 is a step-up transformer that consists of a ferrite core, which has a few turns on the primary and many turns on the secondary. The oscillating (ac) output of Q1 is fed to T1, which, because of its large turns ratio, converts the low-voltage signal into a high-voltage alternating current, which is coupled through resistor R2 to the black-light tube. Resistor R1 and trimmer resistor P1 limit the current flowing through the circuit. As the control on P1 is rotated, more current flows in the circuit, producing a brighter light output.



In A we show two LED output curves derived by experiment. The circuit in B was used to get the data for the short-circuit current plot, while the circuit in C yielded the data for the open-circuit voltage plot.



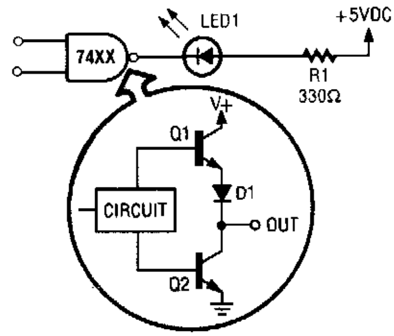
Since LED intensity is linearly related to the input current this circuit can be used to vary the LED's brightness via R2.



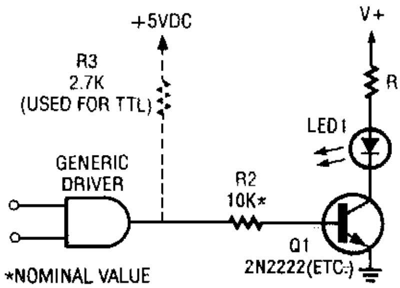
You can drive an LED with an open-collector TTL inverter. The inverter shown must ground the LED to turn it on.

POPULAR ELECTRONICS ***FIG. 52-2***

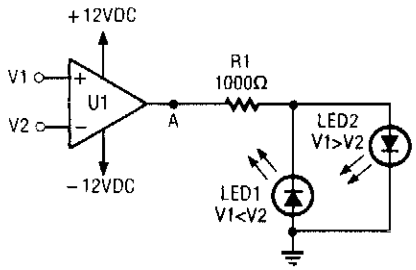
The 12 LED circuits shown are useful for experiments and applications of LED devices. The captions are self-explanatory and illustrate many common LED applications.



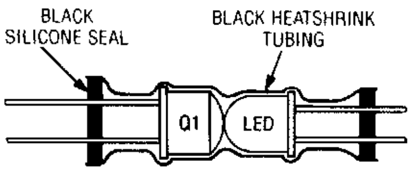
A totem-pole TTL output can drive an LED by grounding the LED's cathode, much like the open-collector driver.



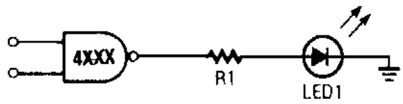
This driver circuit will work for either CMOS or TTL gates, but you don't need R3 in a CMOS-driven circuit.



This is a bipolar output indicator that lets you know if one voltage is greater than, less than, or equal to another.

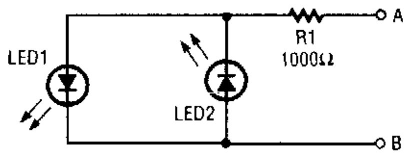


You can "roll your own" optocoupler by using some heat-shrink tubing, an LED, and optical transistor, and silicon sealant as shown here.

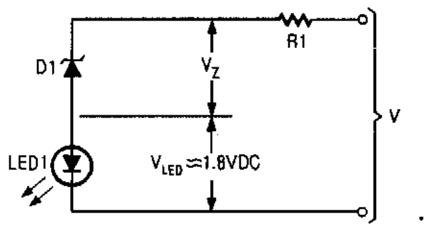


Unlike TTL devices, integrated circuits made with CMOS technology can source enough current to power an LED as shown here.

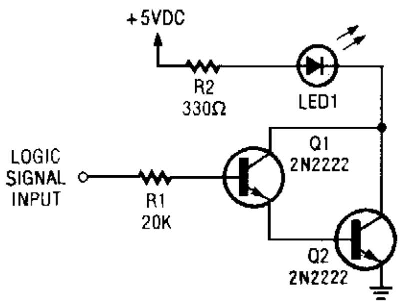
A CMOS-based gate can sink current much like a TTL gate in order to activate an LED.



This simple polarity checker is easy to build and can be of help if you don't know much about a circuit's wiring or grounding convention.



This is a simpler voltage-level sensor than that shown back in Fig. 9. To use it you have to know the polarity of the voltage it is to monitor.



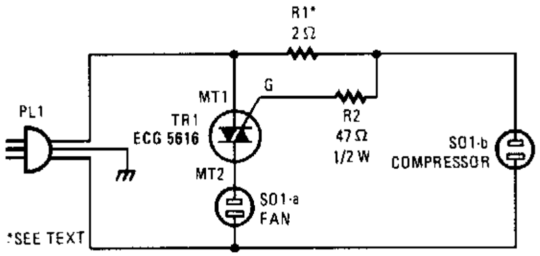
This high sensitivity Darlington LED driver circuit can be used as a simple logic probe. You may have to vary the value of R1 to suit the circuit under test.

53

Load-Sensing Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

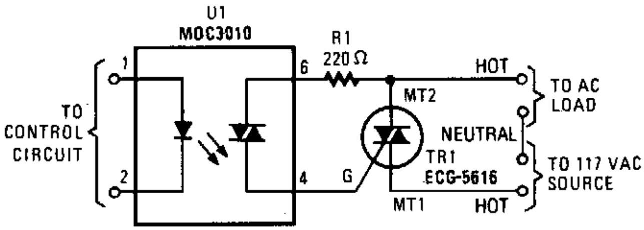
LOAD-SENSING SOLID-STATE SWITCH



When this triac circuit senses current flow through SO1-a, it activates the device plugged into SO1-b. The values of the resistors must be chosen for the specific devices to be plugged in.

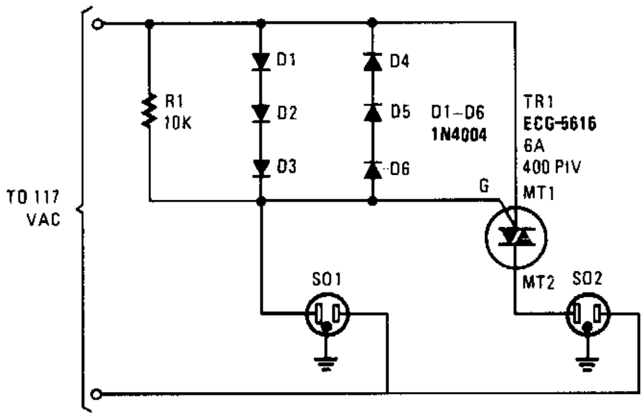
POPULAR ELECTRONICS ***FIG. 53-1***

LOAD-SENSING TRIGGER



Triacs can be controlled by low-power circuits through Triac-driver optoisolators as shown here.

A



B

POPULAR ELECTRONICS ***FIG. 53-2***

A device plugged into SO1 causes a voltage-limited gate trigger for triac TR1, and causes power to be applied to SC2.

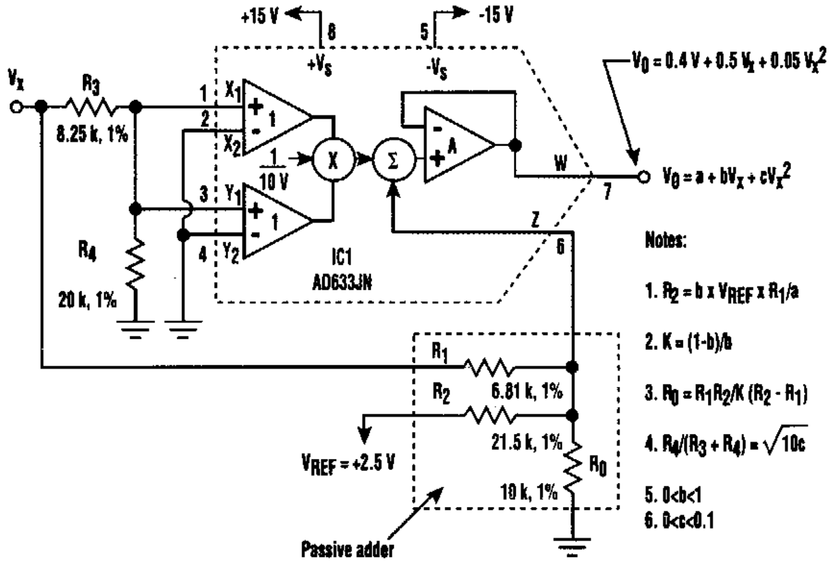
54

Mathematical Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Second-Order Polynomial Generator
Polar-to-Rectangular Converter and Pattern Generator for Radio Direction Finding
Root Extractor

SECOND-ORDER POLYNOMIAL GENERATOR



ELECTRONIC DESIGN *FIG. 54-1*

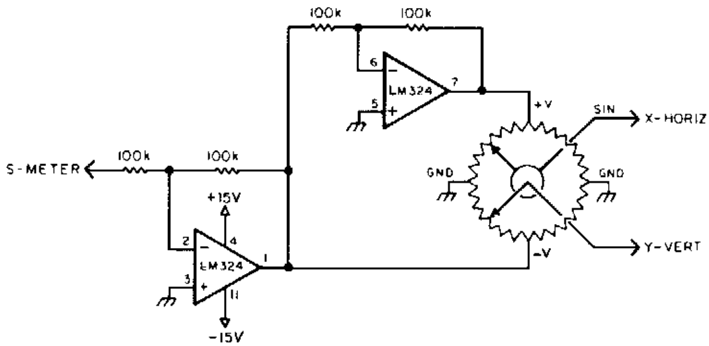
By using a circuit built with a single analog multiplier and five precision resistors, an output voltage (V_o) can be made to create a second-order polynomial.

The circuit implements the following quadratic:

$$V_o = a + bV_x + cV_x^2$$

The input terminals of IC1 are connected to create a positive square term and present the V_x signal to the output with a 1-10-V scale factor. Incorporating the voltage-divider network (resistors R3 and R4) in the input signal path provides additional attenuation adjustment for the coefficient (c) of the square term in the quadratic. Then, the passive adder (resistors R1, R2, and R_o) is wired to IC1's internal summing circuit to generate the polynomial's other two terms; the offset term (a) and the linear coefficient (b).

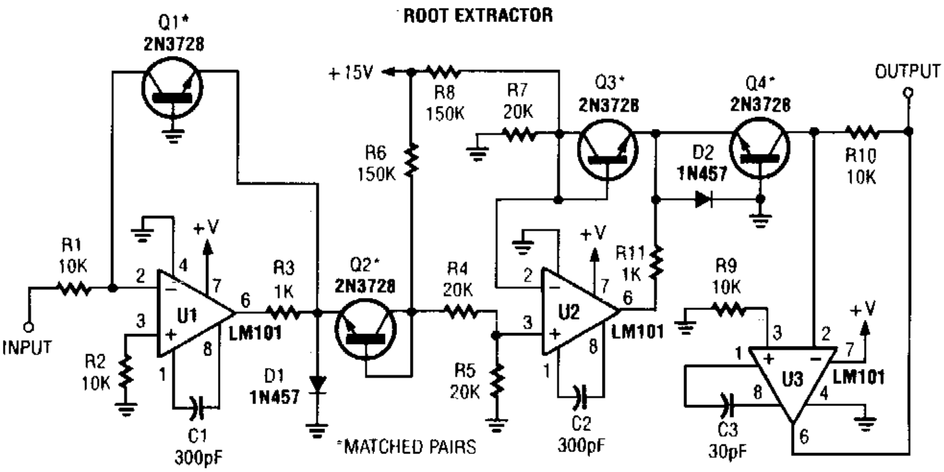
POLAR-TO-RECTANGULAR CONVERTER AND PATTERN GENERATOR FOR RADIO DIRECTION FINDING



73 AMATEUR RADIO TODAY ***FIG. 54-2***

In order to display polar quantities (magnitude and direction of a received radio signal), a sine and cosine voltage proportional to an angle (antenna direction) is needed. In this case, a sine-cosine potentiometer coupled to a directional antenna and a sample of a voltage proportional to received signal is used to display relative magnitude and direction of a received signal.

ROOT EXTRACTOR



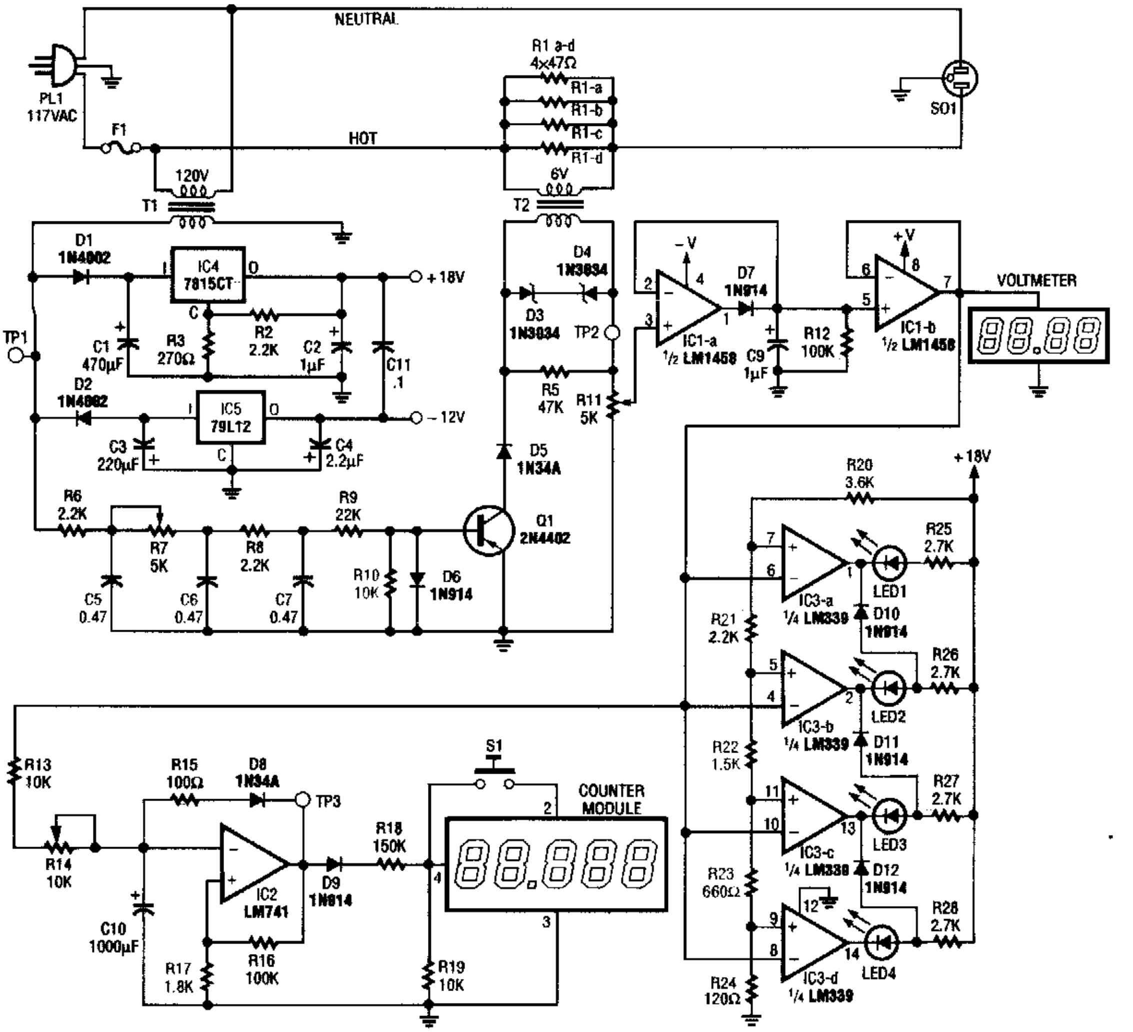
POPULAR ELECTRONICS ***FIG. 54-3***

55

Measuring and Test Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ENERGY CONSUMPTION MONITOR

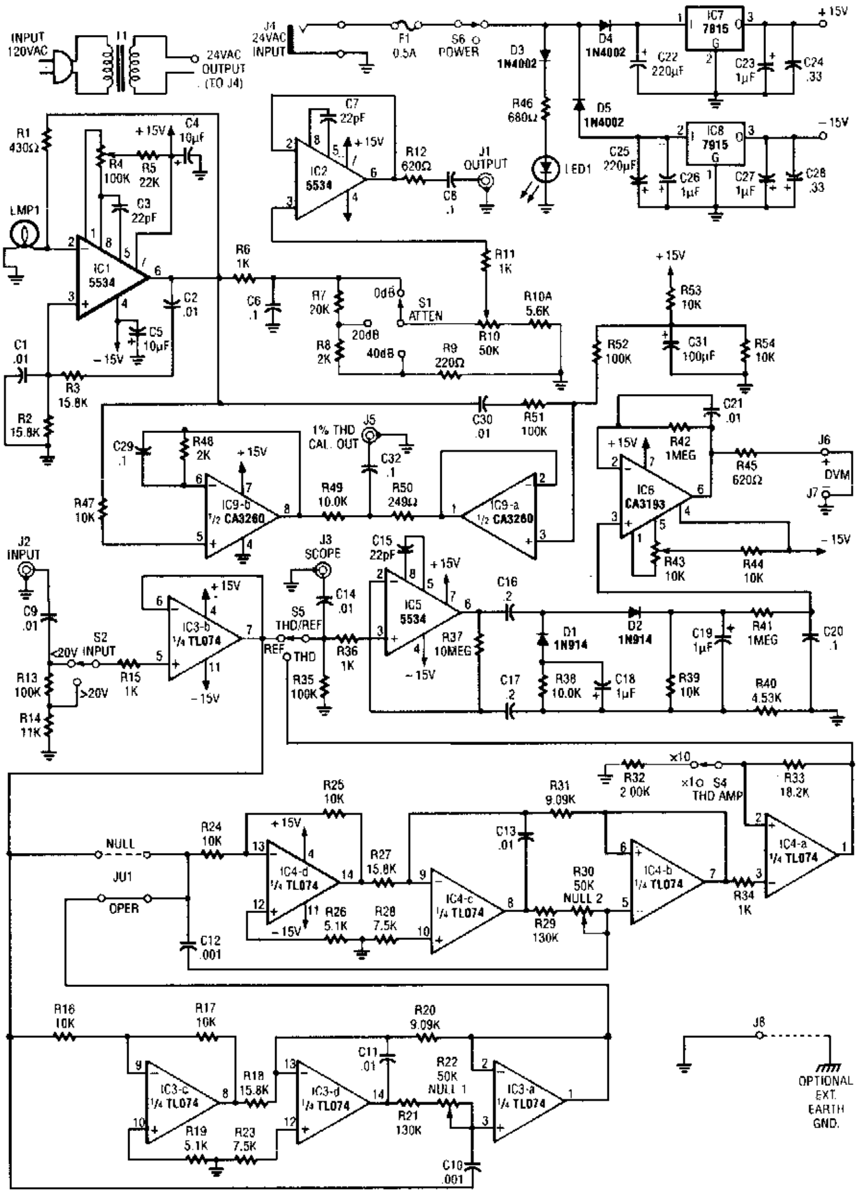


RADIO-ELECTRONICS

FIG. 55-1

The ECM circuit consists of four sections, as shown in the block diagram. A power converter generates a voltage that is proportional to the true of real power consumed by the load. That voltage feeds both a bargraph and a voltage-to-pulse converter. The bargraph gives an approximate indication of the amount of power used, and the voltage-to-pulse converter produces a pulse whose frequency is proportional to the power. The pulse triggers the counter module, which displays the cost of powering the monitored load.

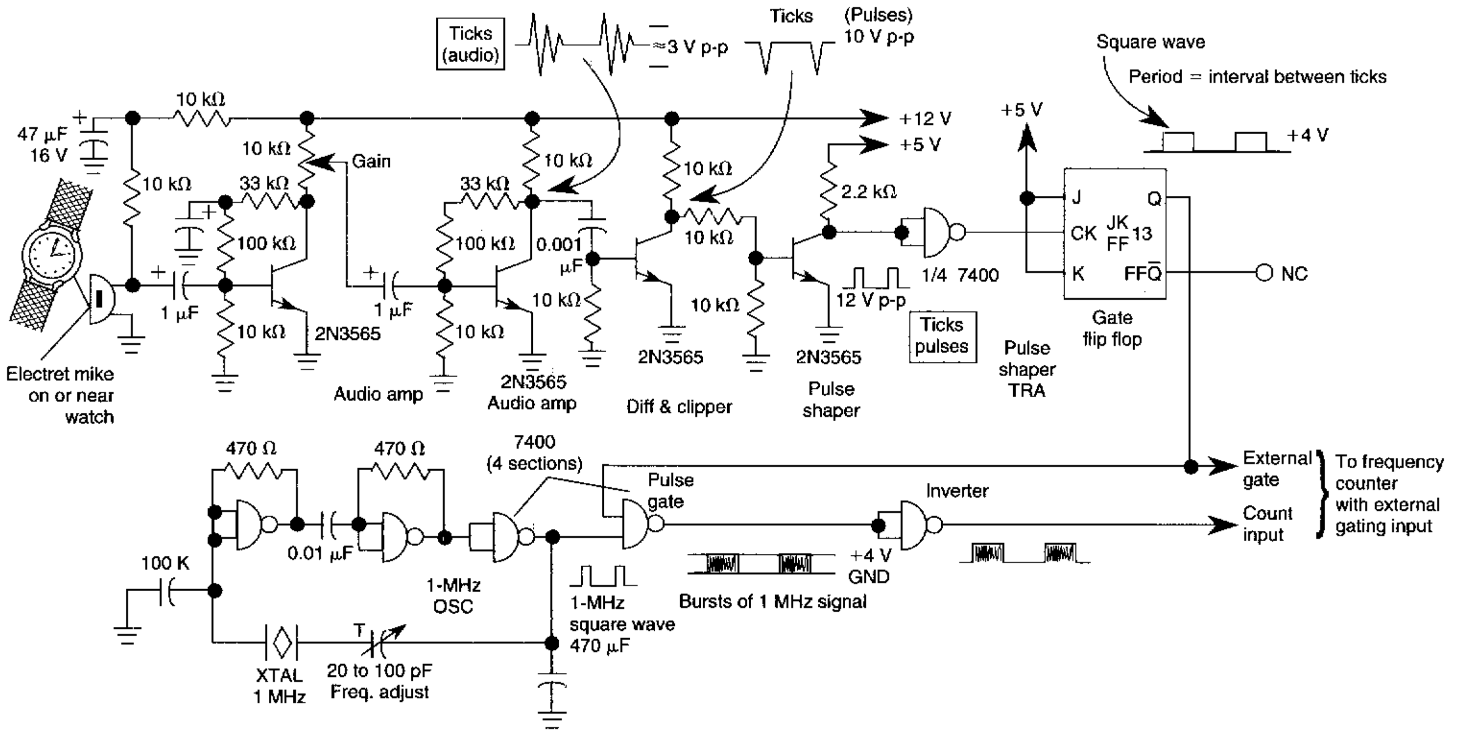
HARMONIC DISTORTION ANALYZER



RADIO-ELECTRONICS *FIG. 55-2*

The circuit includes a low-distortion, 1-kHz oscillator and will measure THD at a user selected voltage level for voltage amplifiers, or for checking amplifiers of power levels to 600 W. It will detect THD levels of .005% (−86 dB). A built-in one-percent THD calibrator is included. The output device is a digital multimeter (DMM).

WATCH TICK TIMER

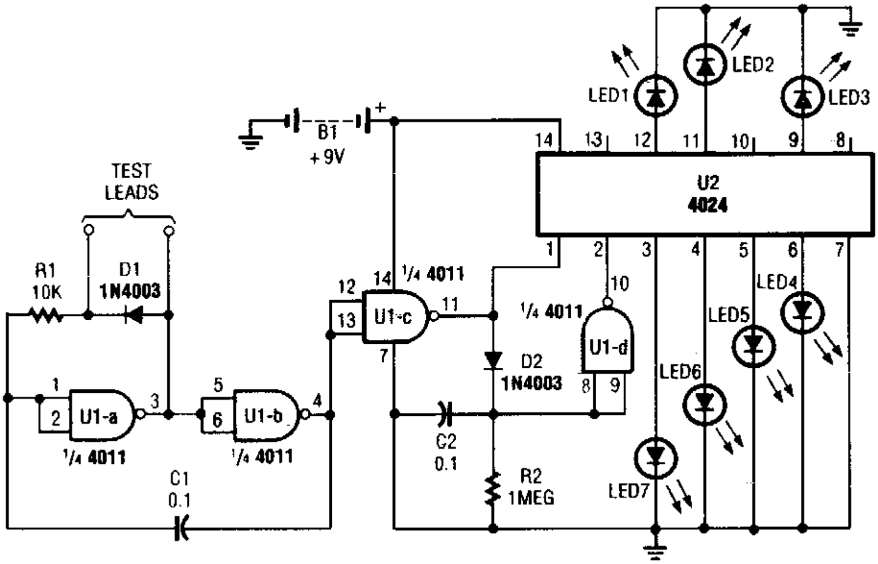


WILLIAM SHEETS

FIG. 55-3

This circuit adapts a frequency counter to measure intervals. It was originally used as a shutter speed checker for a photo application. The watch ticks are clipped and shaped and formed into a square wave. This square wave is used to gate an accurately known clock (1-MHz TTL XTAL OSC) and an external counter is used to directly count the clock pulses during the interval to be measured. A 1-MHz clock can be used to measure to a resolution of 1 µsec. Accuracy = ± time base ±1 µs ±1 count LSB.

VISUAL CONTINUITY TESTER



POPULAR ELECTRONICS

FIG. 55-4

By judging the rate at which a particular LED flashes, you'll be able to estimate the resistance. The circuit consists of two IC's (1 4011 CMOS quad 2-input NAND gate, U1; and a 4024 binary counter, U2), seven LEDs, and a handful of additional components. All of the gates in U1 are wired as inverters.

Two of the inverters (U1-a and U1-b) comprise an astable-multivibrator (free-running oscillator) circuit, whose operating frequency depends on the amount of resistance detected between the test probes. Feedback from the output of the oscillator (at pin 4 of U1-b) back to the input of the circuit (at U1-a, pins 1 and 2) is provided via C1. Resistor, R1, along with the unknown resistance between the test probes, completes the RC timing circuit. The frequency of the oscillator decreases as the resistance between the test probes increases.

The output of the oscillator is fed to pin 12 and 13 of U1-c, the output of which then divides along two paths. In the first path, U1-c's output is applied to the clock input of U2 (a 4024 binary counter) at pin 1; in the other path, the signal is fed through D2 and across capacitor C2, causing it to begin charging. The charge on C2 is applied to U1-d at pins 8 and 9. The output of that inverter (U1-d) is fed to the reset terminal (pin 2) of U2. If there is continuity or a measurable resistance between the test probes, U2's reset terminal is pulled low, triggering the counter and allowing it to process the input pulses (count).

The rate of the count is proportional to the resistance between the test probes. If the resistance between the test probes is low, the counter advances slowly. The counter provides a 7-bit binary output that is wired to seven LEDs.

When the test probes are placed across a short circuit, LED7 flashes. If the tester is placed across a resistance of, for example, 2 MΩ, LED1 will flash. In either case, the LED whose assigned value most closely corresponds to the resistance connected between the two probes will flash continually at a steady pace, while the other LEDs will seem to flash intermittently.

RC DECADE BOX

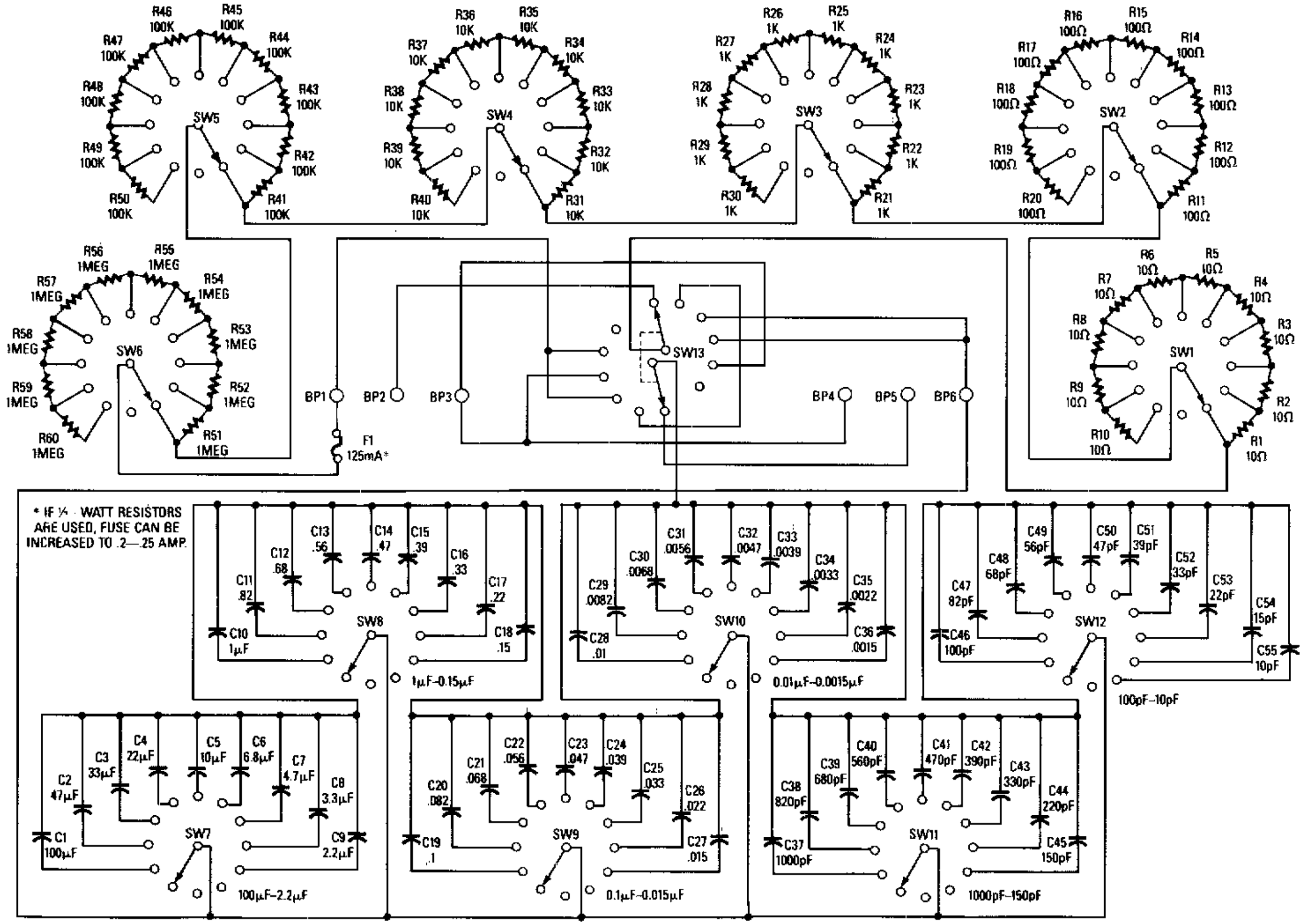
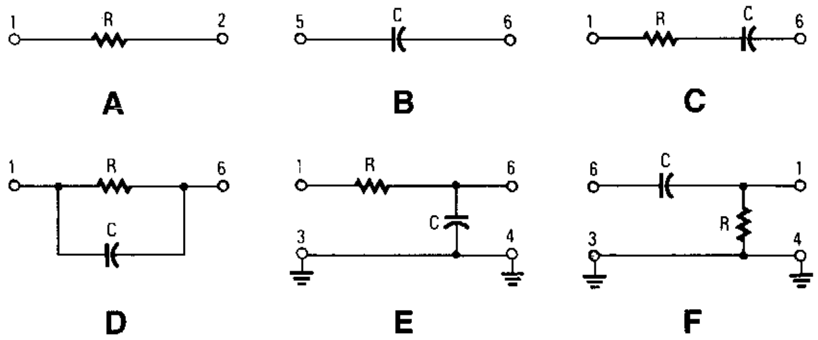


FIG. 55-5



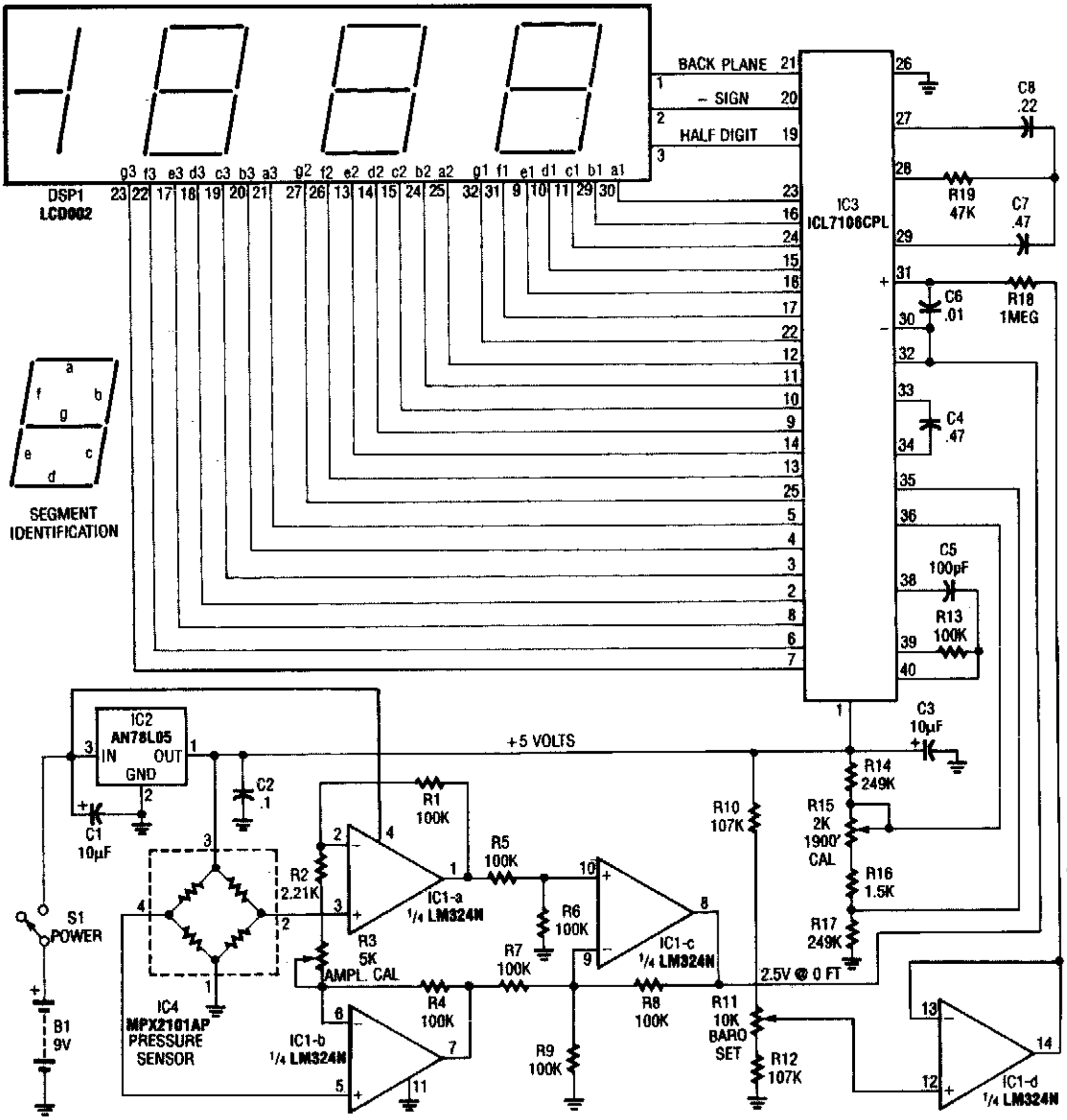
THE VARIOUS CONFIGURATIONS are set using S13: *(a)* resistor only and *(b)* capacitor only (both in position R/C); *(c)* series RC (position SER); *(d)* parallel RC (position PAR); *(e)* Low-Pass Filter (position LPF); and *(f)* High-Pass Filter (position HPF). The terminal numbers listed are those of binding-posts BP1–BP6.

TABLE 1—DECABOX TERMINAL CONNECTIONS

Configuration	S13 Position	IN/GND	OUT/GND
Resistance	R/C	IN: BP1	OUT: BP2
Capacitance	R/C	IN: BP5	OUT: BP6
Series RC	SER	IN: BP1	OUT: BP6
Parallel RC	PAR	IN: BP1	OUT: BP6
Low Pass Filter (Integrator)	LPF	IN: BP1 GND: BP3	OUT: BP6 GND: BP4
High Pass Filter (Differentiator)	HPF	IN: BP6 GND: BP3	OUT: BP1 GND: BP4

This decade box can be set for any resistance value between 10 Ω and 11.1 MΩ in 10-Ω stops. A switch can be used to configure several RC configurations. Use close tolerance components in the circuit. If possible, check components with an accurate bridge or other means to ensure accuracy.

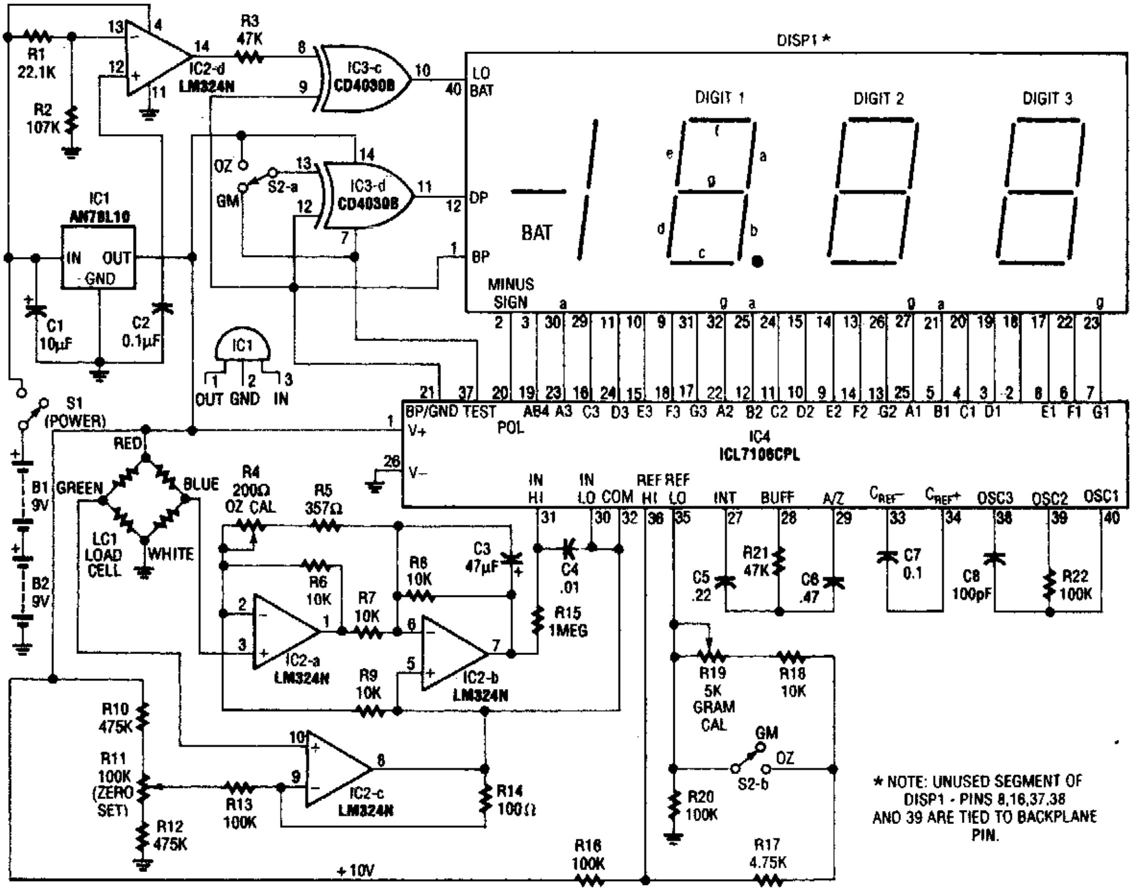
DIGITAL ALTIMETER



RADIO- ELECTRONICS ***FIG. 55-6***

A pressure sensor (IC4) is used with a dc amplifier to convert the bridge output (IC4) to a single-ended voltage. IC1d provides a reference voltage for setting barometric pressure. IC3 is an A/D converter manufactured by Intersil. This drives an LCD module. Calibration reads out in fact. A vacuum pump and a water-based manometer can be used for sensor calibration.

ELECTRONIC SCALE

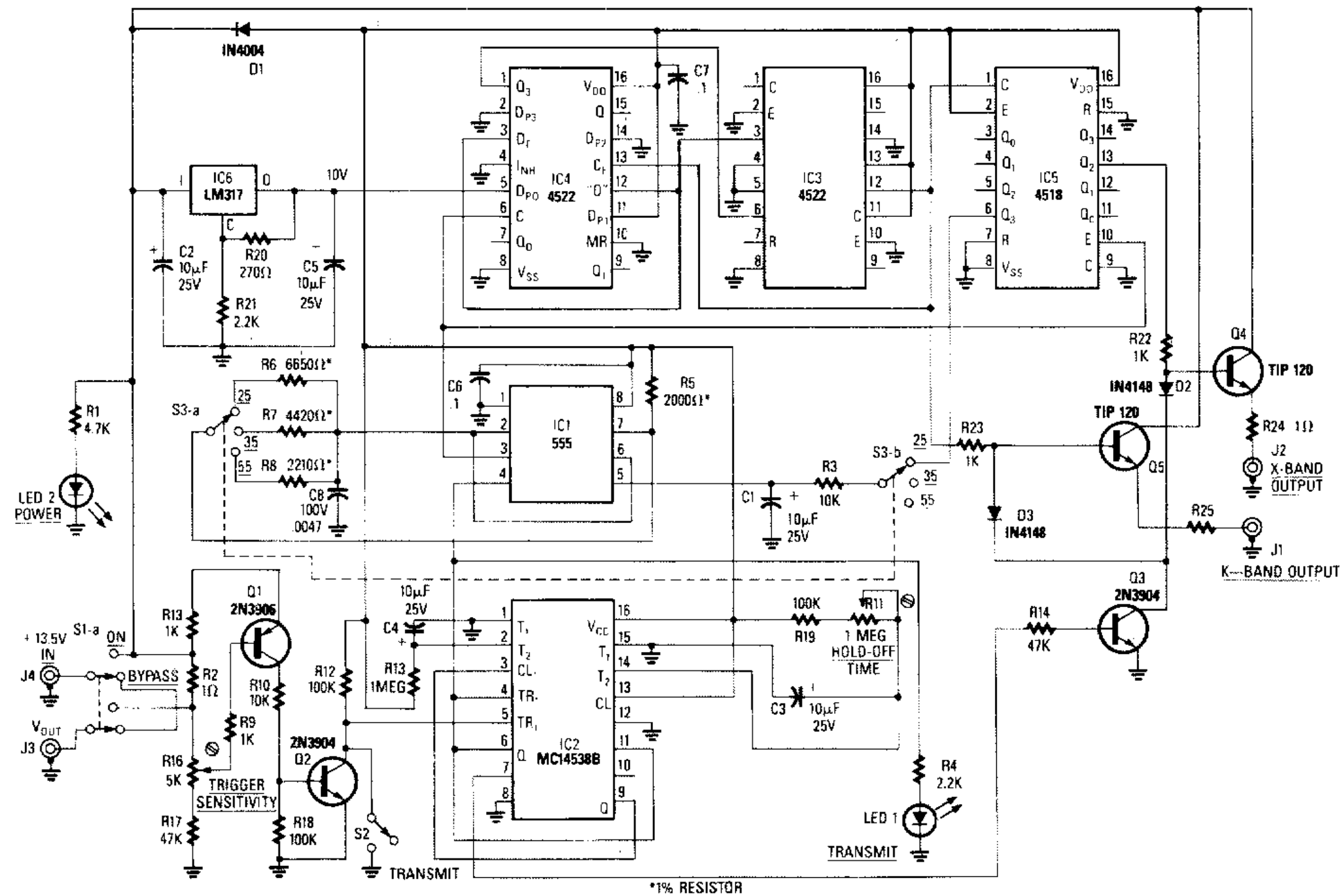


ELECTRONICS NOW

FIG. 55-7

An electronic scale using a pressure transducer (load cell) and an analog-digital (A/D) converter to drive a digital display is shown. The scale range depends on load cell. Display is calibrated in appropriate units. Components are on main circuit and display boards. The off-board controls are on the front panel and case. The cell in this scale is rated for 1.3 pounds (600 grams).

RADAR CALIBRATOR

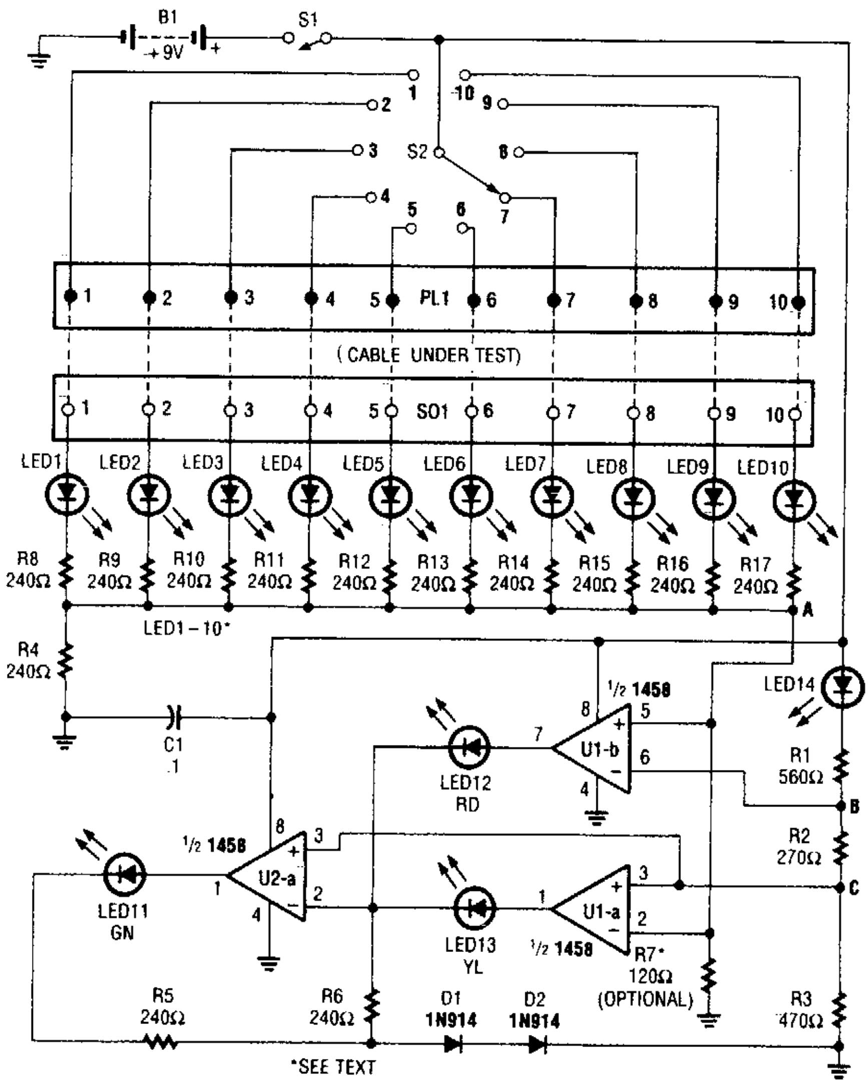


RADIO-ELECTRONICS

FIG. 55-8

This circuit is basically a system that generates a pulsed modulation signal for a Gunn diode microwave oscillator. Several speed settings are preset (S3 a and b). A 555 timer is used with a frequency divider chain to produce Doppler shift equivalents of 25, 35, and 55 mph, for both X- and D-band radars.

CABLE TESTER

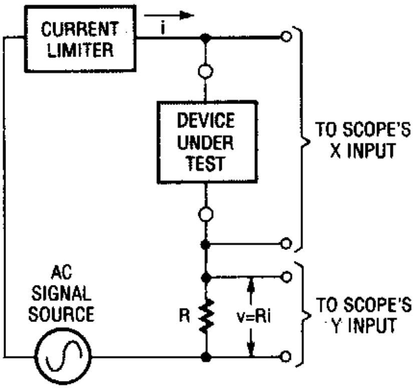


POPULAR ELECTRONICS

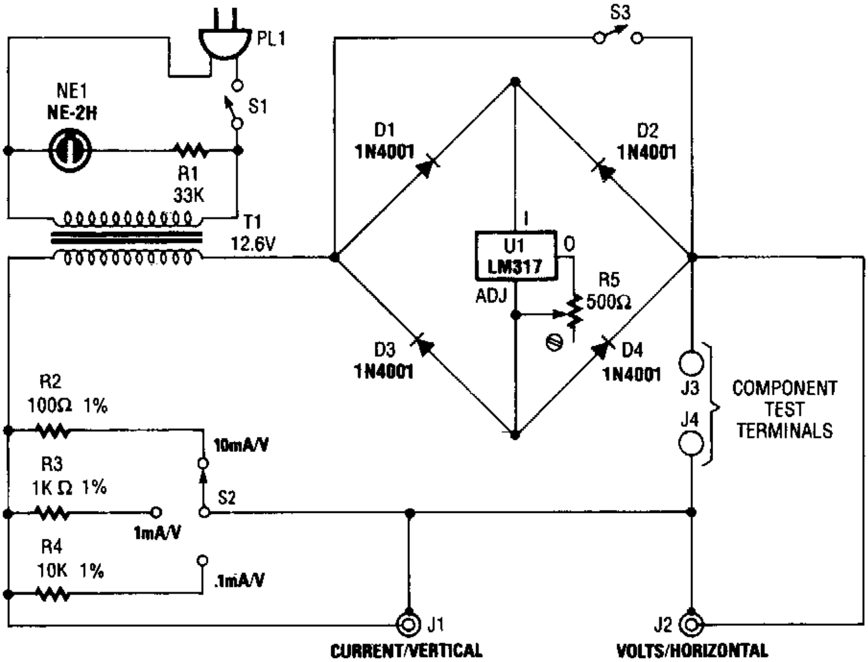
FIG. 55-9

At the heart of the cable tester are two op amps, which are used as a window comparator to indicate a short- or open-circuit condition. A third op-amp comparator is used to indicate a good circuit (i.e., neither open nor shorted). Colored LEDs are used to show the condition of individual conductors within the cable under test; a red one to indicate a short between conductors, a yellow one to identify an open conductor, and a green one to signify that the conductor is okay. Individual LEDs of a bar-graph display are used to show which conductor in the cable is being tested.

SIMPLE CURVE TRACER



This is a simple block diagram of the EZ-Curve. Current-limited AC signals are passed through both the device under test and a precision resistor to yield current and voltage readings.

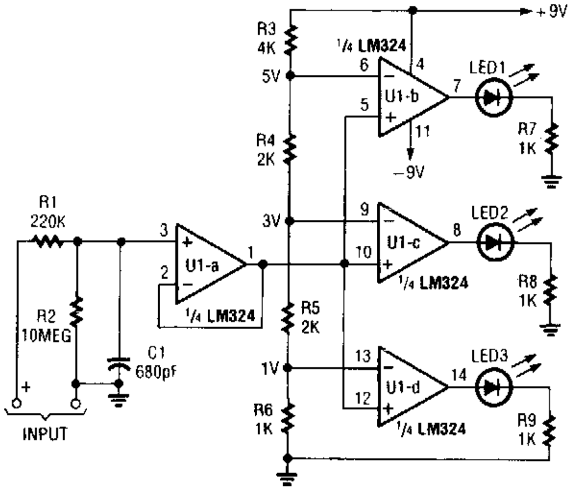


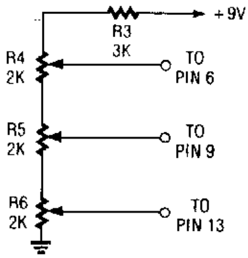
POPULAR ELECTRONICS

FIG. 55-10

Useful for checking diodes, transistors, triacs, SCRs, resistors, and LEDs, this curve tracer should prove useful in the experimenter's lab. It displays the volt-ampere characteristic of a two-terminal device on an oscilloscope.

VOLTAGE LEVEL CIRCUIT





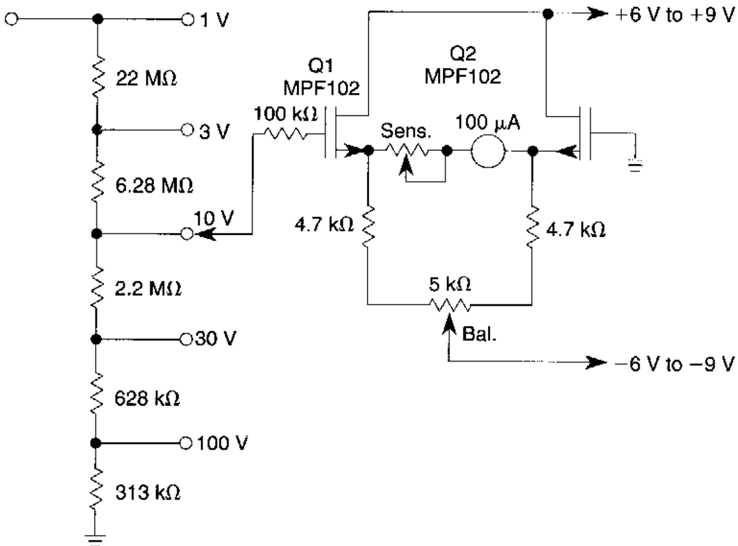
The variable voltage divider shown here can be used with the circuit to allow a wide range of input voltage settings.

B

POPULAR ELECTRONICS **A** ***FIG. 55-11***

A DC op amp and a comparator with a ladder reference divider allow a dc input voltage to light one or more LEDs, depending on voltage levels.

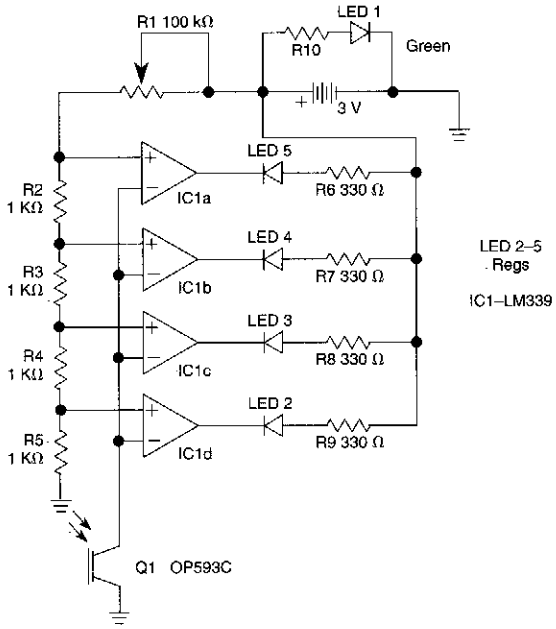
LOW-DRIFT dc VOLTMETER



WILLIAM SHEETS ***FIG. 55-12***

This voltmeter uses a pair of JFETs in a balanced-bridge source-follower amplifier circuit. Q1 and Q2 should be matched within 10% for I_{DSS}. This minimizes meter drift and maintains bridge balance over temperature.

LIGHT METER



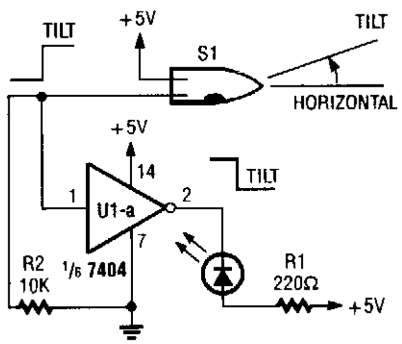
WILLIAM SHEETS

FIG. 55-13

The outputs from the comparators will swing, in sequence, from high to low as the input voltage rises above the reference voltage applied to each comparator. The output LEDs will then switch on in sequence as the voltage rises.

The inverting inputs of the comparators are connected in common to the collector of phototransistor Q1. When Q1 is illuminated, its collector-emitter junction conducts, thereby placing all the inverting inputs within a few millivolts of ground. For most settings of R1, each of the four reference voltages exceeds the value. Therefore, when Q1 is illuminated, the output from each comparator is high and its respective indicator LED is off.

MERCURY SWITCH TILT DETECTOR

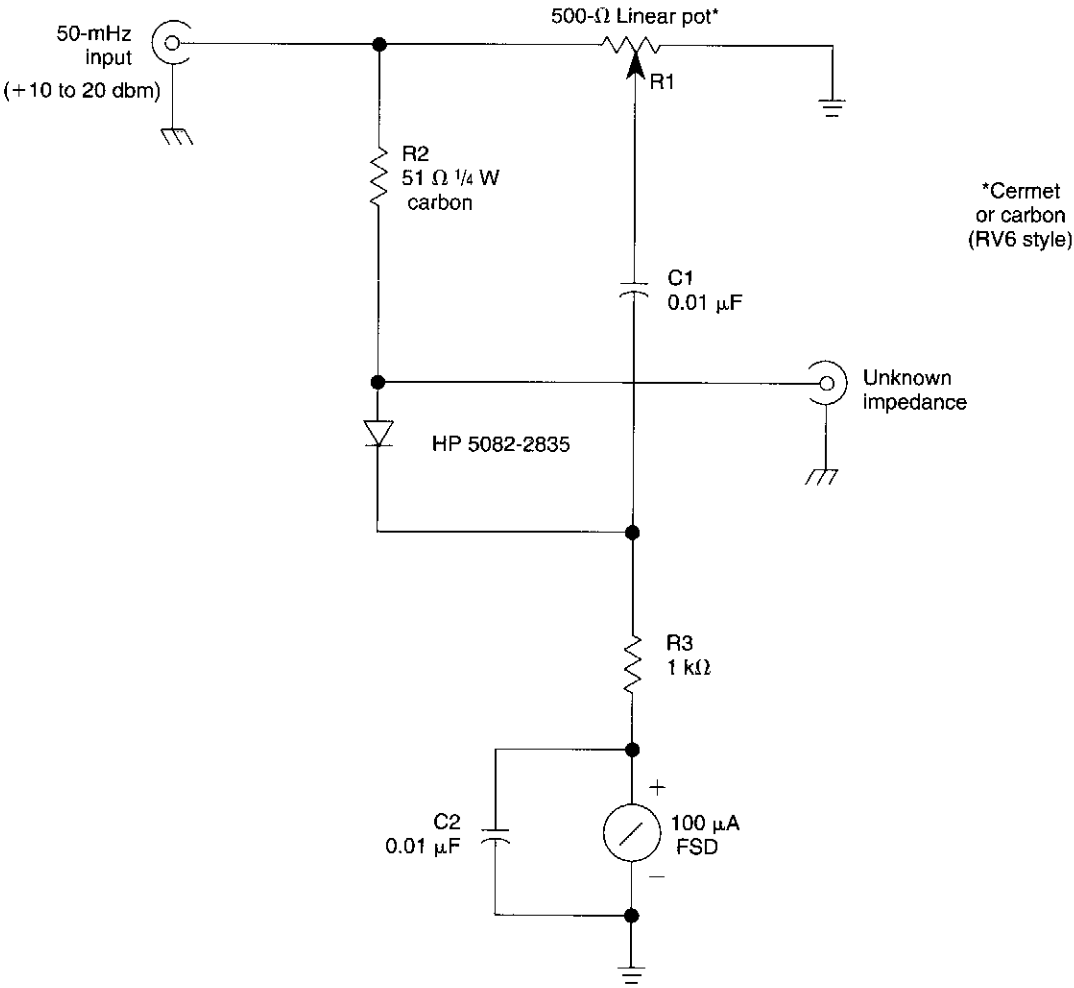


POPULAR ELECTRONICS

FIG. 55-14

If the mercury bulb in this circuit is tipped, U1-a will light LED1 by going low, indicating a "tilted" condition.

50-MHz RF BRIDGE

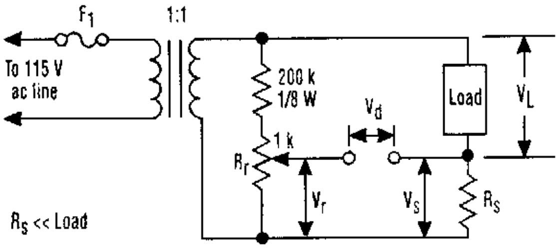


WILLIAM SHEETS

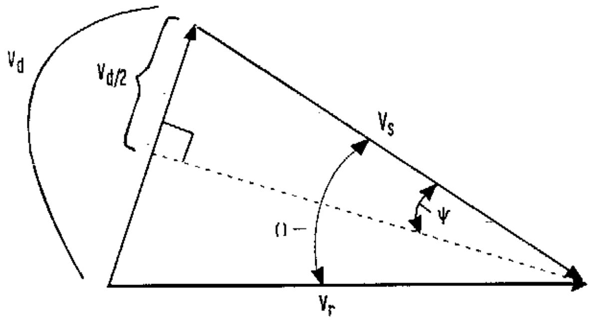
FIG. 55-15

The bridge shown was used for measurements on 50-MHz amateur radio antennas. R1 is a miniature 500 Ω linear potentiometer. The unknown impedance is compared to R2, a 51-Ω resistor. An external signal source is required.

ac WATTS CALCULATOR



The load's power factor, which is the cosine of the phase angle between the voltage across the load current, can be calculated simply with this circuit. A 1:1 isolation transformer is used to prevent direct contact with the line



By properly adjusting R_r, the vector diagram of voltages V_s, V_d, and V_r forms an isosceles triangle, which simplifies the power calculation.

ELECTRONIC DESIGN

FIG. 55-16

The method basically consists of determining the power factor of the load—the cosine of the phase angle between the voltage across the load and the load circuit. Using a simple circuit, that angle can be calculated quite simply.

This circuit uses a 1:1 isolation transformer to prevent direct contact with the line. It is wise to proceed with caution whenever voltages of this magnitude are utilized in a test setup, even though the voltages that will be measured are usually below 1 V.

R_s is a circuit-sense resistor and R_r is a multi-turn potentiometer. The voltage across R_r is approximately 0.5% of the line voltage, which should be sufficient for most applications.

R_r is adjusted so that $|V_r| = |V_s|$; then V_d is measured. In the vector diagram according to Kirchhoff's voltage law, V_s, V_d, and V_r form a triangle, which becomes isosceles by adjusting R_r. V_s is in phase with the load current and V_r is essentially in phase with the load voltage.

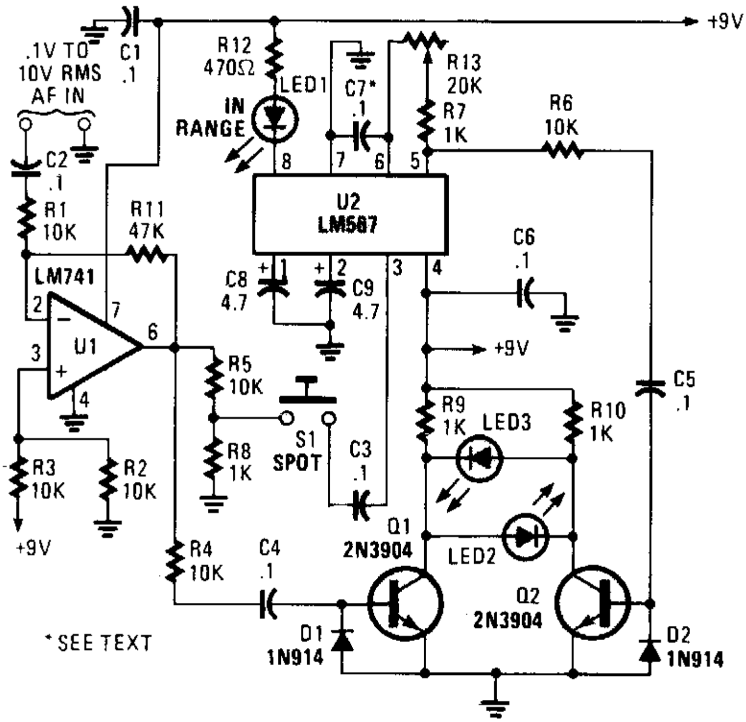
The power delivered to the load can be calculated as follows:

$$P_L = V_L \times I_L \times \text{Cos}\ \theta$$

$$= V_L \times (V_s/R_s) \times \text{Cos}\ [2\ \text{Sin}-1\ (V_d/2V_s)]$$

$$[\theta\ 2\ \psi = 2\ \text{Sin}-1\ (V_d/2V_s)]$$

AUDIO-FREQUENCY METER



POPULAR ELECTRONICS

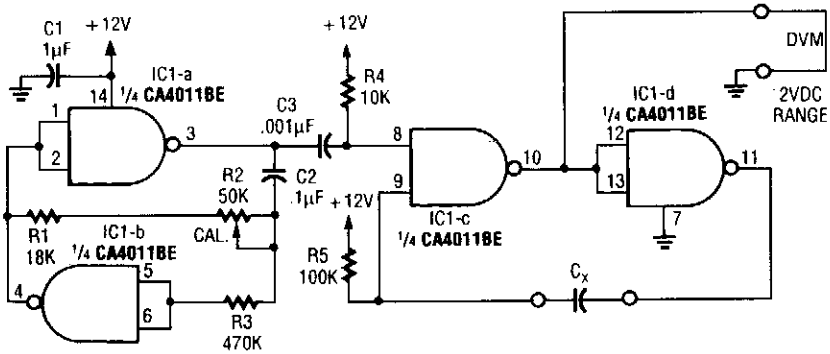
FIG. 55-17

This meter differs from the norm in that it does not use a D'Arsonval movement or digital display to give a reading of the input frequency. Instead, the measured frequency is read from a hand-calibrated dial.

Any audio signal applied to the circuit is amplified by U1 and the resulting output is divided along two paths. In one path, the output signal is applied to the mixer; in the other path, the signal is applied to the input of U2 through S1 (a normally open pushbutton switch).

The portion of the amplifier signal that is fed to the mixer is applied to the base of Q1, causing it to toggle on and off at the signal frequency. In the other path, when S1 is pressed, a portion of the op amp's output is applied to U2. If the signal is within the range of U2's internal oscillator's operating frequency, LED1 lights, and a signal is fed to the base of Q2. If the two signals arriving at the mixer do not match exactly, LED2 and LED3 light. That means that the circuit must be fine tuned, which is accomplished by releasing S1 and fine tuning R13 until LED2 and LED3 go out. The dial setting at that point gives the frequency of the input signal to within 1 Hz (or as close as the calibrated dial will allow).

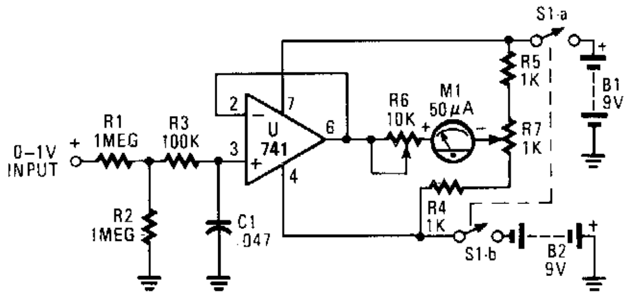
ONE-IC CAPACITANCE TESTER



RADIO-ELECTRONICS **FIG. 55-18**

This circuit can be used to match capacitors, etc. The dc output voltage is related to the capacitance values of C_X. The circuit values shown are for capacitors in the 0.01-µF order of magnitude, but they can be changed for lower or higher values.

TRANSISTOR CHECKER

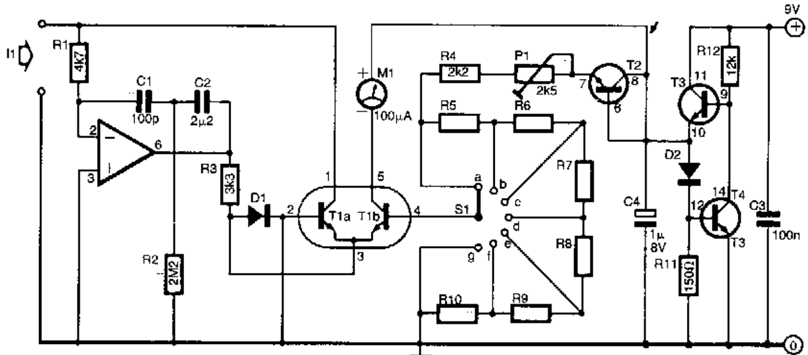


POPULAR ELECTRONICS **FIG. 55-19**

The circuit is built around a 741 general-purpose op amp that is configured as a voltage follower; with the components shown, the op amp has a voltage gain of one. The output of the 741 is used to drive a 50-µA meter movement. Potentiometer R7 is used to zero the meter and R6 sets the meter's full-scale reading.

Calibrating the meter is a snap. With no input applied to the circuit, set R6 to mid-position and adjust R7 to zero the meter. Once that is done, apply a positive 1-Vdc voltage to the input and adjust R6 for a full-scale reading. The voltmeter can be adjusted to read both positive and negative voltages by adjusting R7 for a center scale reading at the meter's zero position and a positive 1-V reading at the meter's full-scale position.

LOW-CURRENT AMMETER



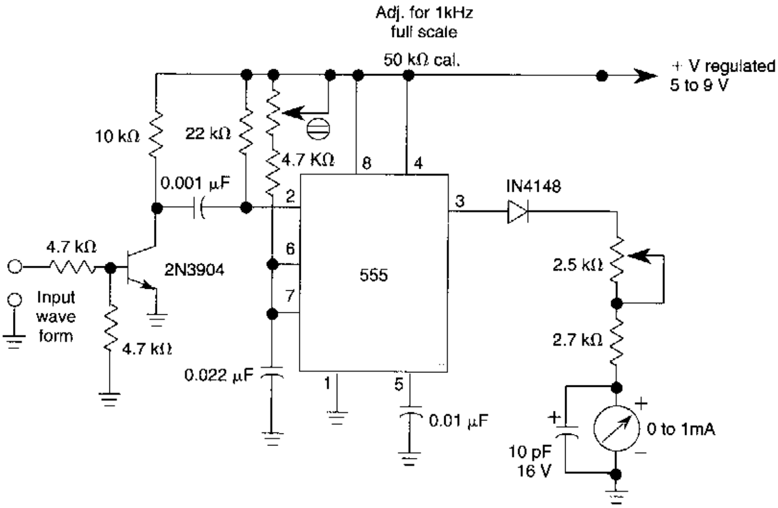
T1. . .T4 = IC1 = CA3046
A1 = IC2 = LF411, OP41
R5. . .R10 = 511Ω, 1%
D1. . .D2 = 1N4148

S1	I1
a	100 pA
b	1 nA
c	10 nA
d	100 nA
e	1 μA
f	10 μA
g	100 μA

303 CIRCUITS ***FIG. 55-20***

Without using high-value precision resistors, this circuit uses a current mirror, T1a/T1b. Currents of 100 pA can be measured with this circuit. M1 is a 100-mA meter. Make sure to use a high-quality PC board and low-leakage circuit construction.

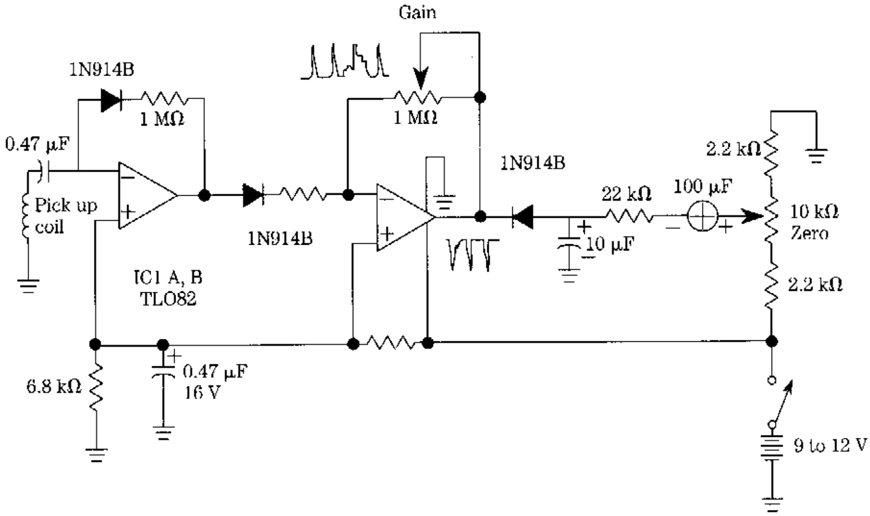
ANALOG FREQUENCY METER



WILLIAM SHEETS ***FIG. 55-21***

This 1-kHz linear-scale analog frequency meter circuit uses the 555 as a pulse counter. Frequency is read on M1, (or 1 mA meter) which can be calibrated to read 0 to 1 kHz.

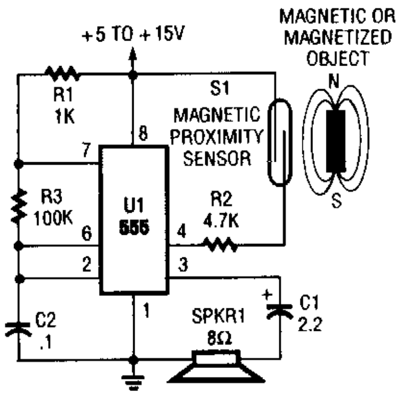
ELECTROMAGNETIC FIELD SENSOR



WILLIAM SHEETS *FIG. 55-22*

A telephone pick-up coil is used as a sensing coil. Any 60-Hz hum picked up by the sensing coil is rectified, amplified, and detected, and then drives a meter.

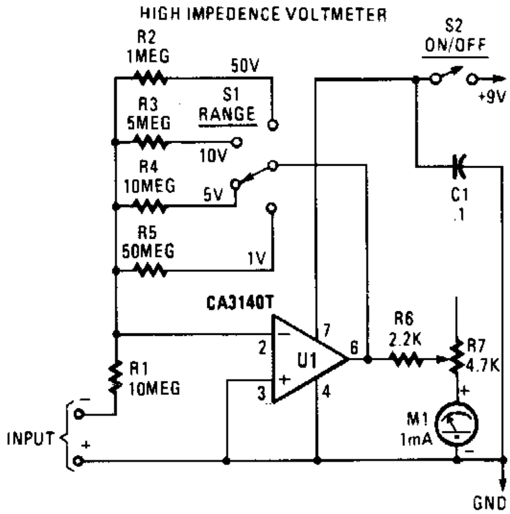
MAGNETIC PROXIMITY SENSOR



POPULAR ELECTRONICS *FIG. 55-23*

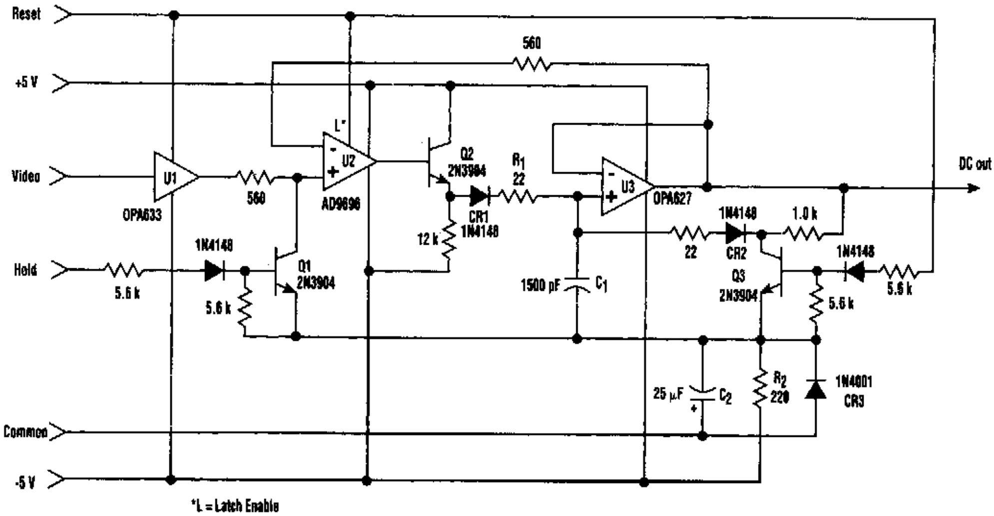
A magnetic need switch enables a 555 oscillator, which drives a speaker. C2 can be varied for different tone frequencies.

HIGH-IMPEDANCE VOLTMETER



POPULAR ELECTRONICS *FIG. 55-24*

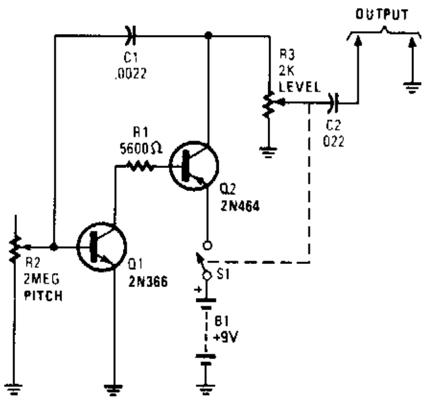
FAST VIDEO SIGNAL AMPLITUDE MEASURER



ELECTRONIC DESIGN *FIG. 55-25*

Video-signal amplitude can be measured with this simple circuit, which is basically a modified standard peak detector. The device can verify RGB generated by video RAMDACs. U1 is a high-speed buffer and U2 is a latched comparator. C1 is a hold capacitor. Reset is performed by Q3. U2 has a latch that maintains the last comparator state. The reset holds the comparator output low during the reset operation. The dc output voltage is equal to the signal's maximum amplitude.

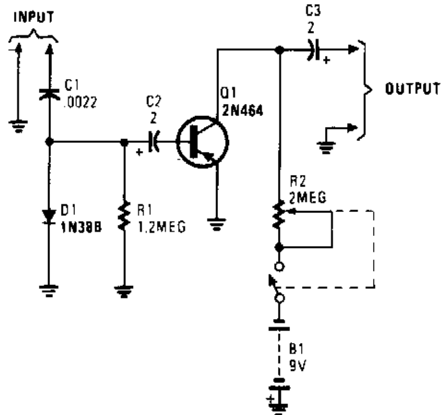
SIGNAL GENERATOR



POPULAR ELECTRONICS *FIG. 55-26*

Useful for troubleshooting audio, video, and lower frequency RF amplifiers, this circuit generates a signal that is rich in harmonics.

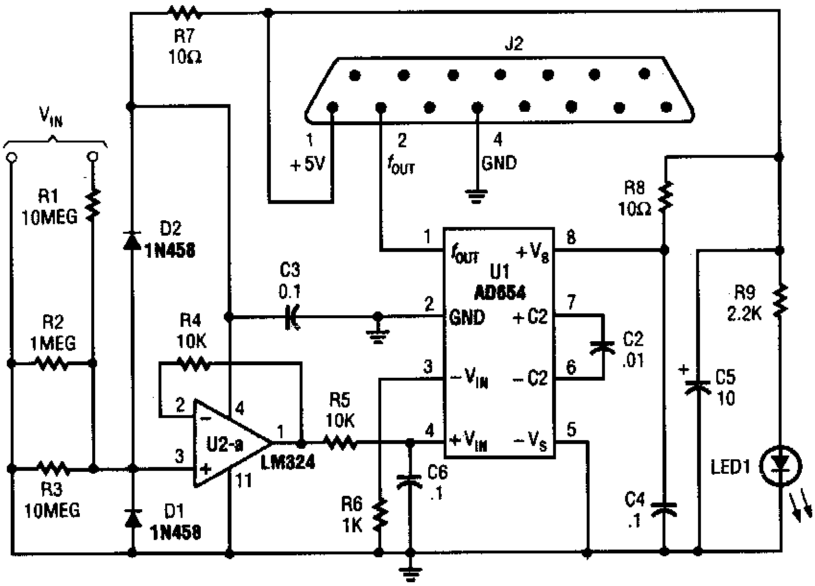
SIMPLE SIGNAL TRACER



POPULAR ELECTRONICS *FIG. 55-27*

In this circuit, C1/D1/R1 form an envelope detector. C2 couples audio to the base of Q1. R2 can be adjusted for the desired gain.

DVM ADAPTER FOR PC

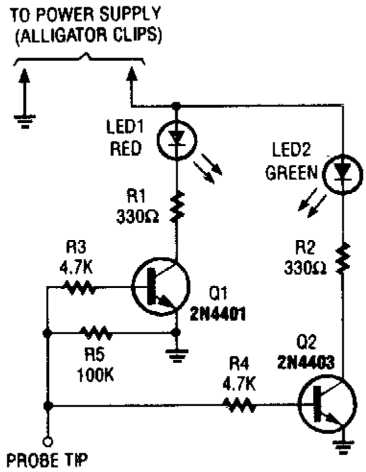


POPULAR ELECTRONICS

FIG. 55-28

The adapter consists of a voltage to frequency adapter with a signal conditioner and protection circuit. J2 connects to the game port of a PC. See reference listed for software for use with this circuit.

SIMPLE DIGITAL LOGIC PROBE

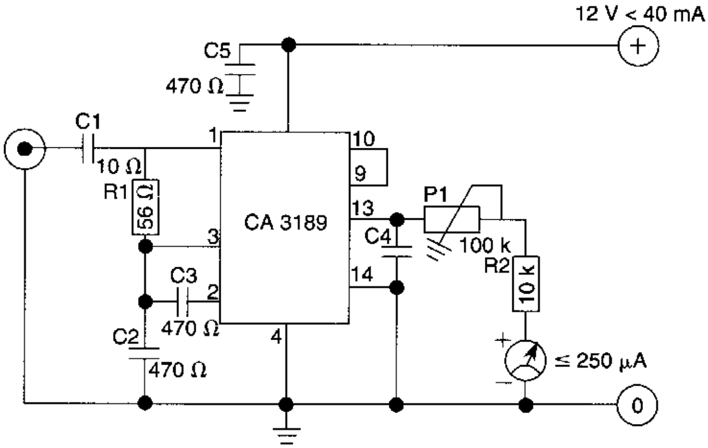


The design of the digital logic probe centers around a pair of complementary bipolar transistors, which, in this application, are used as electronic switches.

POPULAR ELECTRONICS

FIG. 55-29

S METER FOR COMMUNICATIONS RECEIVERS

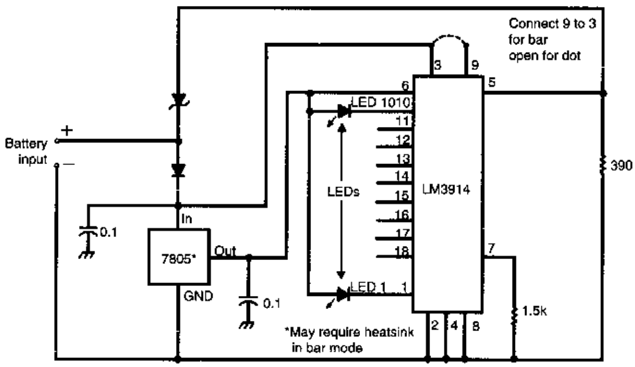


303 CIRCUITS ***FIG. 55-30***

Because many amateur receivers are fitted with an S meter that functions far from logarithmically, the proposed circuit should be a welcome extension of such receivers. Although ICs such as the CA3089 or the CA3189 are not in common use anymore, they serve a useful purpose in the meter circuit, because, apart from a symmetric limiter, a coincidence detector, and an AFC amplifier, they contain a very good logarithmic amplifier-detector.

As is seen, the circuit is fairly simple, but remember that these ICs operate up to about 30 MHz; the wiring of the meter and its connections in the receiver should be kept as short as possible.

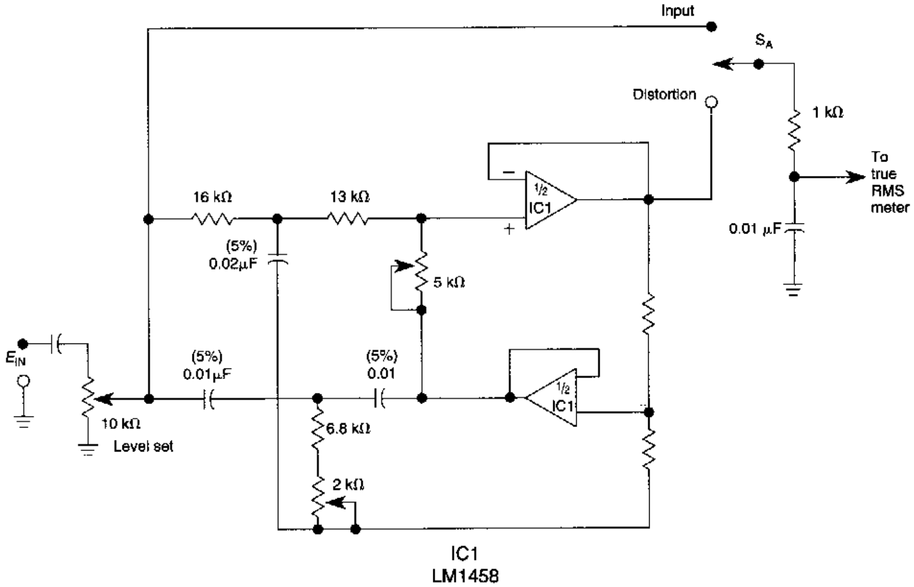
LED EXPANDED SCALE VOLTMETER



73 AMATEUR RADIO TODAY ***FIG. 55-31***

A 10-V zener diode is used to expand the scale of a 0- to 5-V voltmeter to a 10- to 15-V voltmeter. The LED bar graph lights one segment per 0.5-V input above 10 V. The 7805 IC provides a 5-V reference and 5 V for the bar graph LEDs.

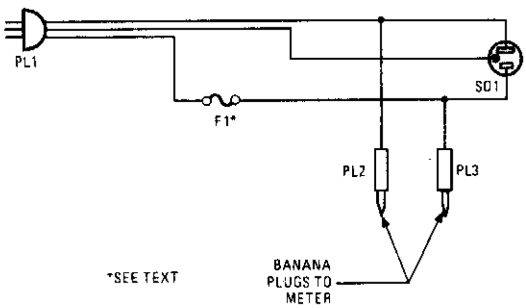
1-KHz HARMONIC DISTORTION METER



WILLIAM SHEETS ***FIG. 55-32***

The circuit useful for distortion measurements notches out the fundamental frequency of 1 kHz to allow measurement of the residual level of harmonics. First a true RMS meter is used to measure the 1-kHz input level E_{in} by setting S_A to the input position. Then, S_A is placed in the distortion position and the 2 k potentiometer is adjusted for a null. The residual reading is noted. The THD is then calculated based on the formula:

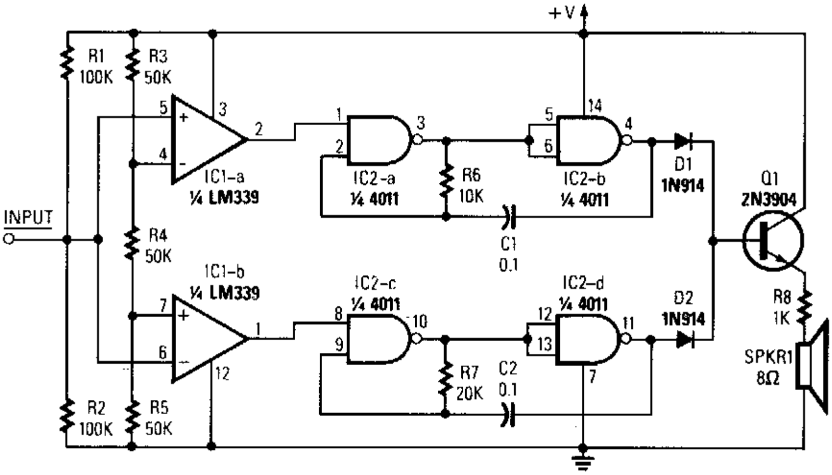
LINE VOLTAGE-TO-MULTIMETER ADAPTER



POPULAR ELECTRONICS ***FIG. 55-33***

This ac line-to-multimeter adapter can make checking line voltage safer. You can use it to find taxing loads on your household wiring.

AUDIBLE LOGIC TESTER



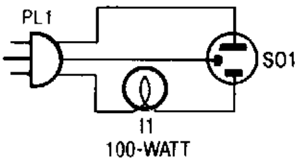
1989 R-E EXPERIMENTERS HANDBOOK *FIG. 55-34*

The tester provides an audible indication of the logic level of the signal presented to its input. A logic high is indicated by a high tone, a logic low is indicated by a low tone, and oscillation is indicated by an alternating tone. The input is high impedance, so it will not load down the circuit under test. It can be used to troubleshoot TTL or CMOS logic.

The input section determines whether the logic level is high or low, and enables the appropriate tone generator; it consists of two sections of an LM339 quad comparator. One of the comparators (IC1-a) goes high when the input voltage exceeds 67% of the supply voltage. The other comparator goes high when the input drops below 33% of the supply. Resistors R1 and R2 ensure that neither comparator goes high when the input is floating or between the threshold levels.

The tone generators consist of two gated astable multivibrators. The generator built around IC2-a and IC2-b produces the high tone. The one built around IC2-c and IC2-d produces the low tone. Two diodes, D1 and D2, isolate the tone-generator outputs. Transistor Q1 is used to drive a low-impedance speaker.

SHORT TESTER FOR 120-V EQUIPMENT



Do you deal with old equipment in unknown condition? If so, this little circuit could keep you from causing further harm to already shorted devices.

POPULAR ELECTRONICS *FIG. 55-35*

DIGITAL PRESSURE GAUGE

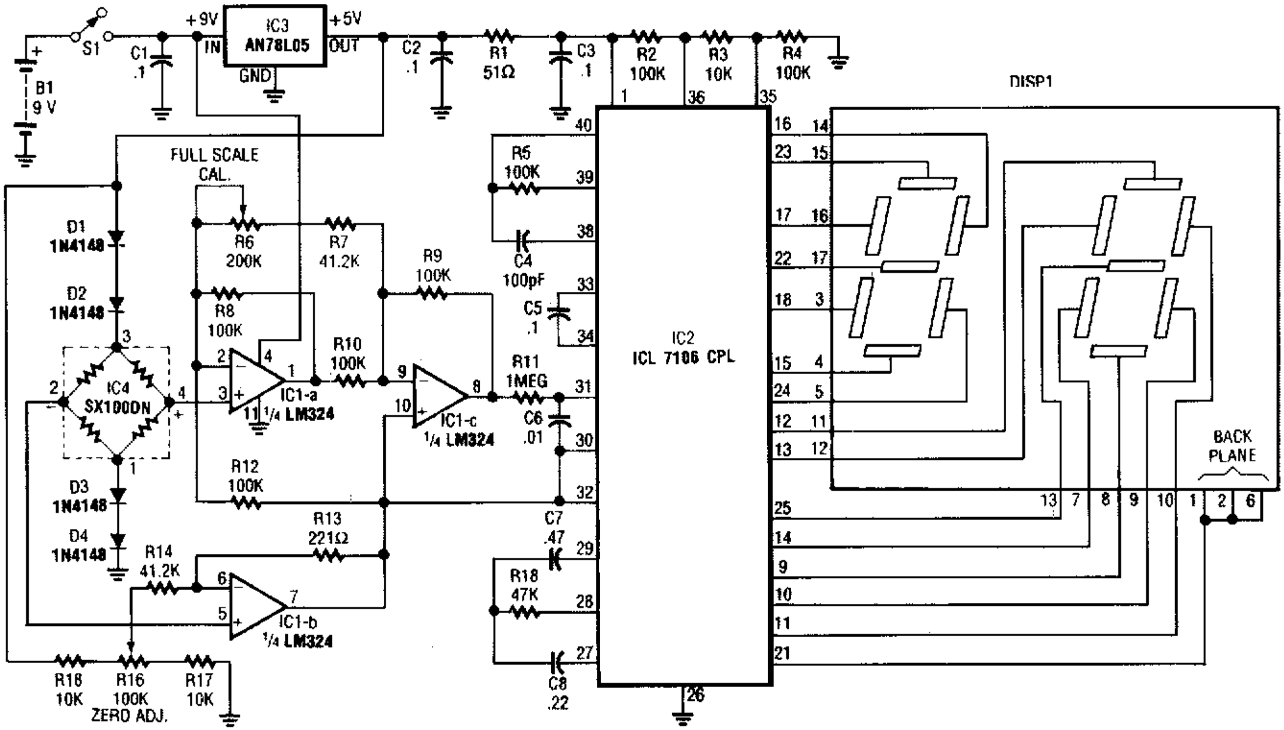
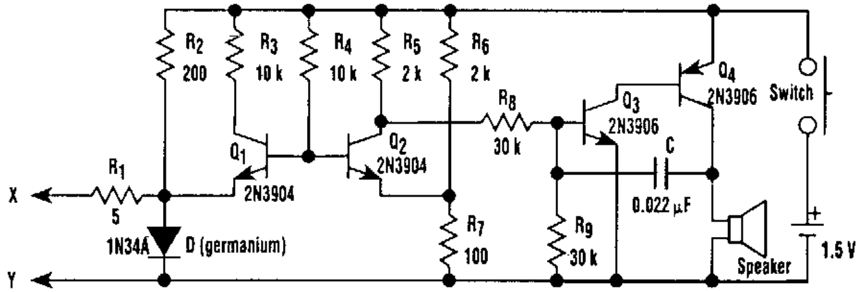


FIG. 55-36

This electronic pressure gauge uses a Wheatstone bridge-type pressure sensor tb drive a 3½ digit A/D converter and a display. IC1 is a pump (quad) that interfaces the bridge sensor to the A/D converter. R16 provides zero adjustment and R6 provides full-scale calibration. D1 thru D4 provide temperature compensation.

SIMPLE SHORT FINDER



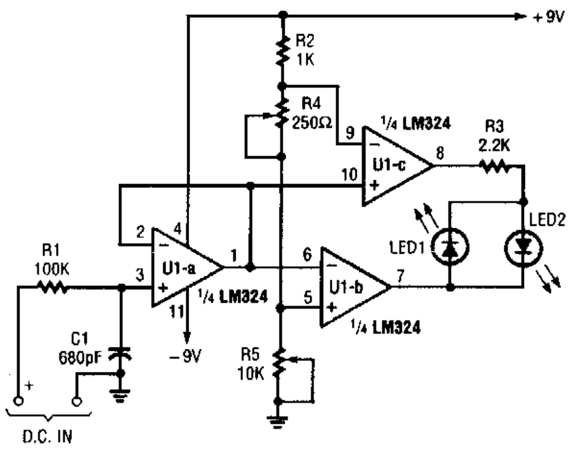
ELECTRONIC DESIGN

FIG. 55-37

Transistors Q1 and Q2, together with resistors R1 through R7, make up the input balancing stage, which senses the resistance between points X and Y. The input stage is essentially a bridge, consisting of R1, R2, R6, R7, and the resistance between points X and Y.

Transistors Q3 and Q4 and their associated passive components form a buzzer, which sounds when the tester detects a short. The buzzer is controlled by the output from Q2. When the input resistance is high (more than about 10 Ω), Q2 turns on, so its collector potential is close to ground, and the buzzer remains off. When the input resistance is sufficiently low, Q2 turns off, and the buzzer sounds. The frequency of the sound, which is about 1000 Hz, can be adjusted by varying the value of capacitor (C).

VOLTAGE MONITOR



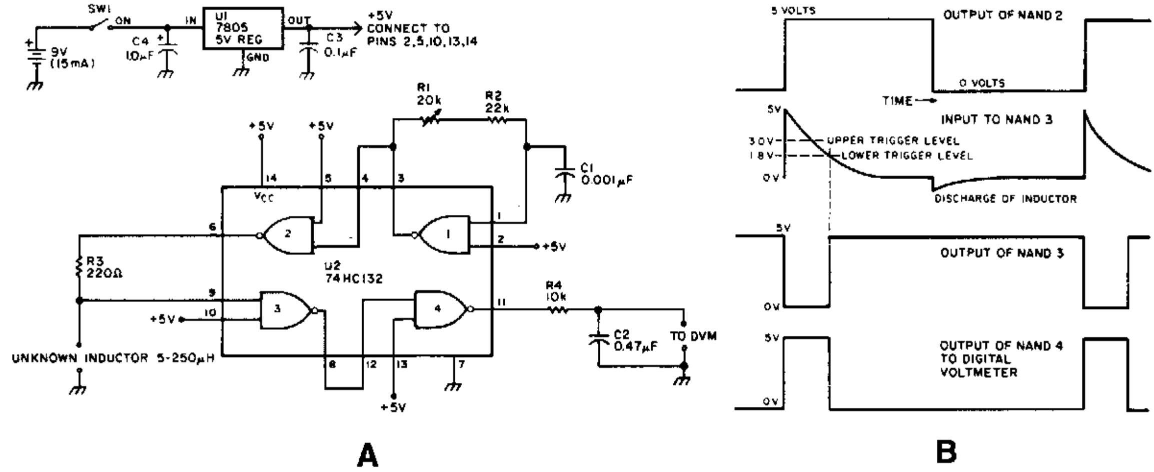
The adjustable voltage monitor can be used to check whether the voltage in a circuit remains within a given range.

POPULAR ELECTRONICS

FIG. 55-38

If the dc voltage is less than the voltage at pin 5 of U1-B, then LED 1 will light. If the voltage is over 5V, LED2 will light. If the voltage is within the window set by R4 and R5, neither LED will light. This circuit is useful as an under-or-over voltage monitor.

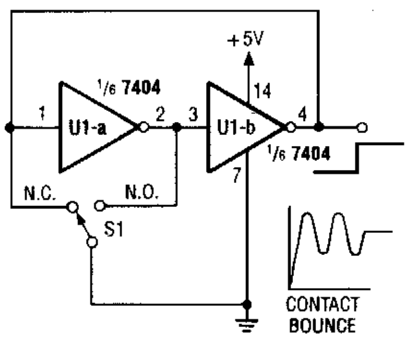
LINEAR INDUCTANCE METER



73 AMATEUR RADIO TODAY *FIG. 55-39*

Using the fact that in an RL circuit, the pulse width seen across the inductor is proportional to the inductance, this circuit reads this indirectly on a DVM. The range is about 5 to 250 μH.

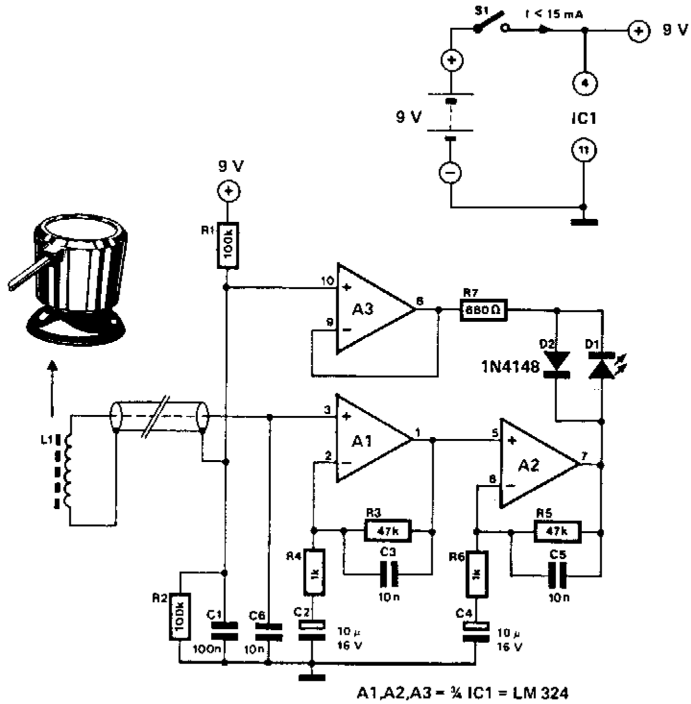
DEBOUNCE CIRCUIT



POPULAR ELECTRONICS *FIG. 55-40*

This debounce circuit will keep the electrical noise generated by the mechanical switch (S1) from reaching the next circuit in line.

ac WIRING LOCATOR

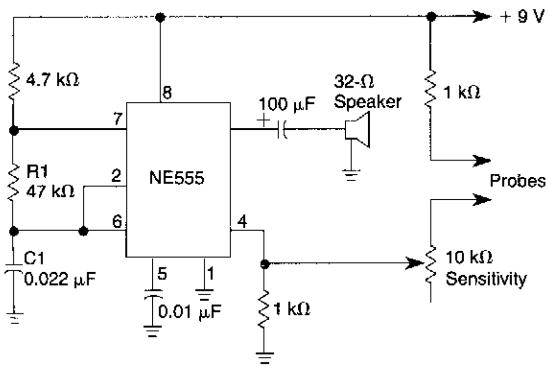


303 CIRCUITS

FIG. 55-41

This circuit uses a pick-up coil to sense the 50- or 60-Hz field around wiring carrying ac. L1 is a telephone pick-up coil with a suction pad. D1 (LED) lights during positive half waves, indicating that ac current is present.

AUDIBLE CONTINUITY TESTER

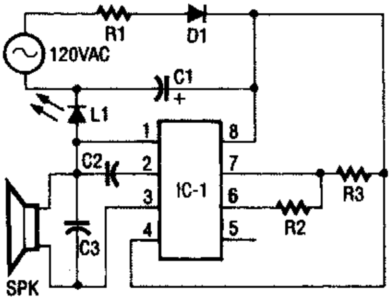


This 555 oscillator sounds a tone when continuity exists between the probes. Oscillator frequency is determined by the values of R1 and C1.

WILLIAM SHEETS

FIG. 55-42

ac OUTLET TESTER

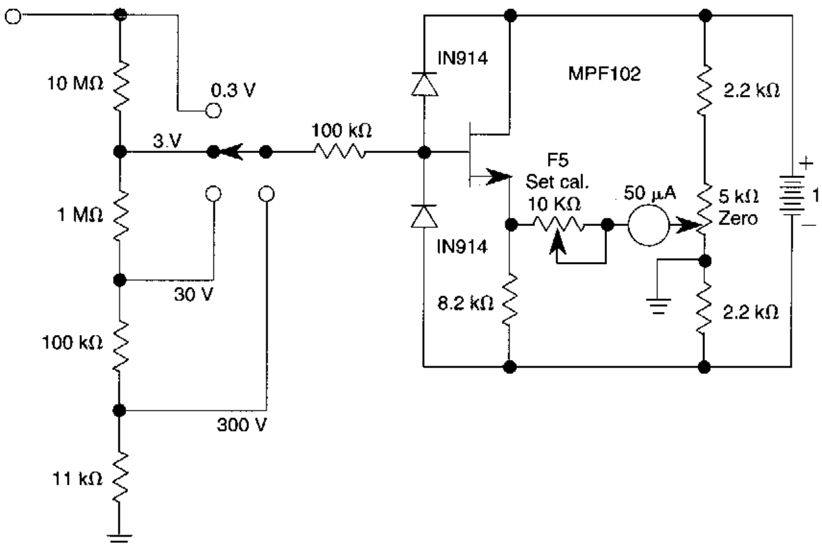


C1 .... 50 μF Electrolytic Capacitor
C2,C3 .. .047 μF Disc Capacitor
D1 .... 1N4003 Diode
IC1 .... 555 Timer IC
L1 .... Jumbo Red LED
R1 ....3.9K, 1 watt Resistor
R2 ....2K, 1/4 watt Resistor
R3 .... 4.7K, 1/4 watt Resistor
SPK .... Piezoelectric Speaker

1991 PE HOBBYIST HANDBOOK ***FIG. 55-43***

The tester consists of a rectifier circuit and a multivibrator circuit. The ac voltage is half-wave rectified by diode D1 and stored in capacitor C1. Resistor R1 is used to limit the current through D1 to a safe value. The voltage stored across C1 supplies IC1 operating power. The IC, the versatile 555 timer, is configured to operate as a multivibration whose operating frequency is determined by C2, R2, and R3. The output of IC1, on pin 3, is coupled to a piezoelectric speaker (SPK), which gives an indication of the presence of ac. An LED (L1) also lights when ac is present.

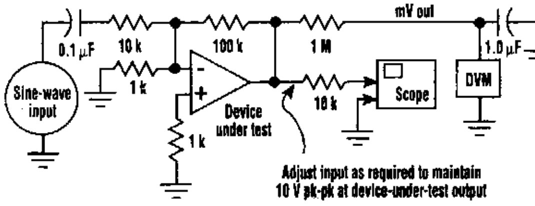
JFET VOLTMETER



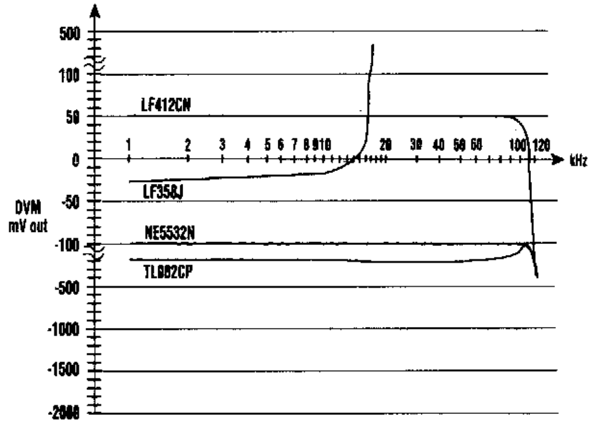
WILLIAM SHEETS ***FIG. 55-44***

This very simple voltmeter circuit uses a 50-μA meter in a bridge circuit. It is useful for noncritical applications.

CHECK FOR OP-AMP dc OFFSET SHIFT



LF412CN		LF358J		NE5532N		TL082CP	
kHz	mV out	kHz	mV out	kHz	mV out	kHz	mV out
1	51	1	-23	1	-101	1	-313
10	50	10	-14	10	-101	10	-314
100	44	12	-9	100	-109	50	-301
110	9	13	-4	110	-152	60	-296
115	-358	14	4	120	-378	70	-279
120	-502	15	7			80	-258
130	-1374	16	58			90	-227
140	-1741	17	119			100	-184
		17.5	156			110	-125
		18	494				



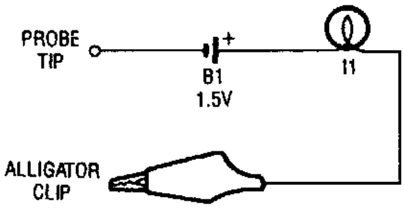
ELECTRONIC DESIGN

FIG. 55-45

The dc values of op-amp offsets can't always be taken for granted when delivering ac outputs. No device is ever exactly symmetrical for maximum positive slew rate versus maximum negative slew rate. Consequently, there is always some range of output slew rates in which the device used limits in one direction more severely than in the other. What results in rectification of the ac signal and an apparent shift of the dc offset.

This test circuit can check for the shift phenomenon. The accompanying table and graph illustrate the results obtained for four devices, all of different types. As frequency and slew rate are increased, the effect can be either relatively abrupt (LF412CN and NE55532N) or relatively gradual (LF358J and TL082CP).

CONTINUITY TESTER FOR LOW-RESISTANCE CIRCUITS

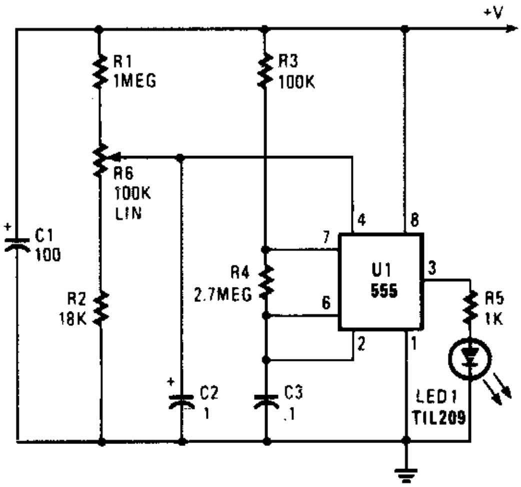


POPULAR ELECTRONICS

FIG. 55-46

The continuity tester is little more than a battery and a lamp connected in series, with one end of the string terminated in an alligator clip, and the other end connected to the probe tip.

SUPPLY VOLTAGE MONITOR

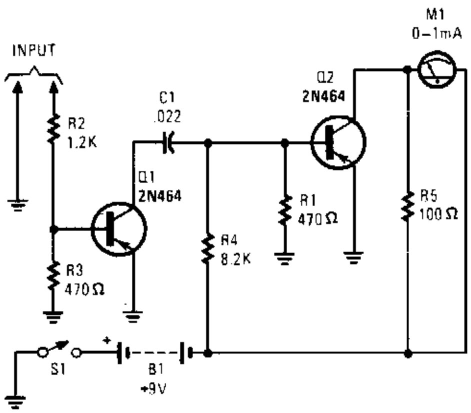


Excessive voltage causes U1 to oscillate, causing LED1 to flash. R6 sets the desired trip level.

POPULAR ELECTRONICS

FIG. 55-47

AUDIO-FREQUENCY METER CIRCUIT

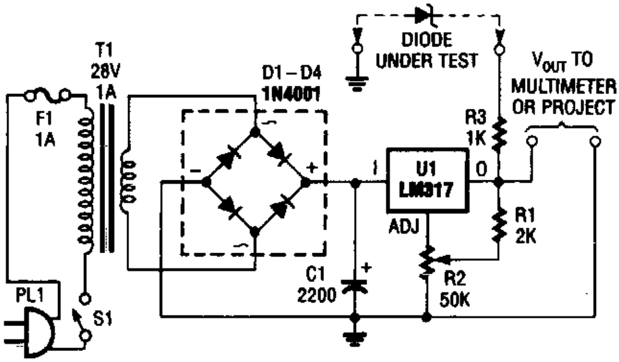


This simple tachometer circuit uses a pulse shaper Q1 to drive M1, a 0- to 1-μA meter. C1 can be varied to optimize operation.

POPULAR ELECTRONICS

FIG. 55-48

ZENER DIODE TEST SET



POPULAR ELECTRONICS *FIG. 55-49*

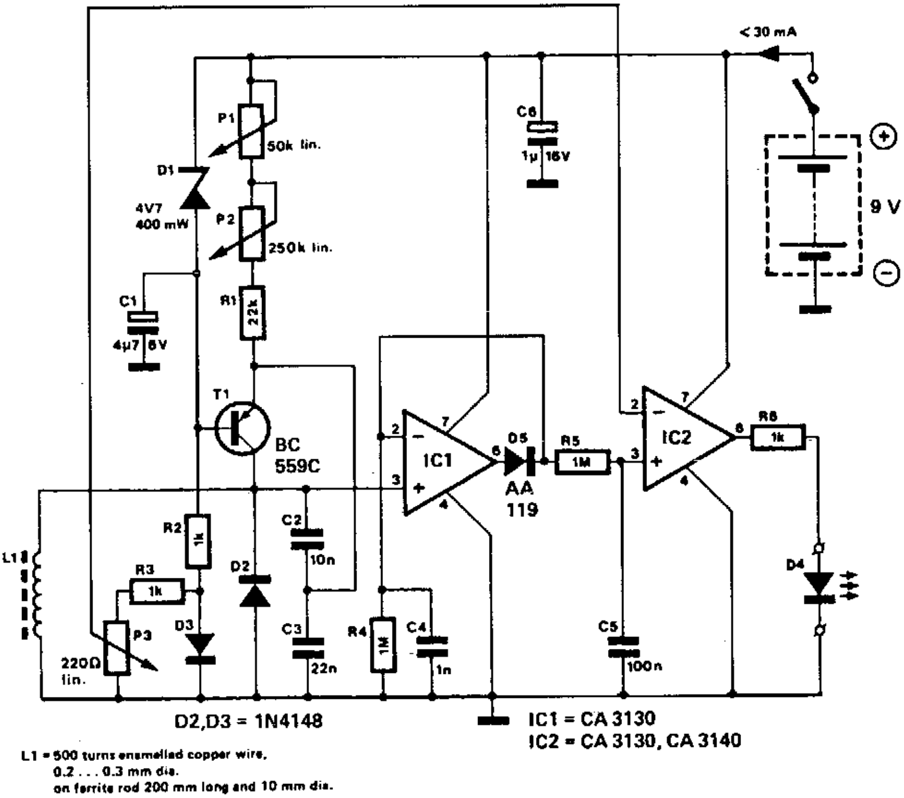
This versatile circuit can be used to test zener diodes or act as a stand-alone power supply. It requires a voltmeter to work as a zener tester.

56

Metal-Detector Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

METAL PIPE DETECTOR

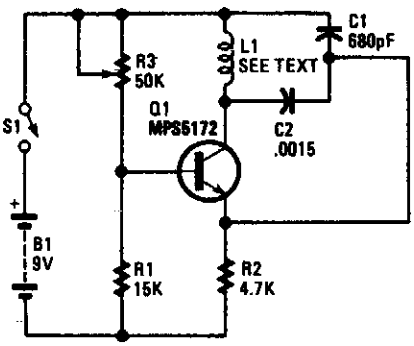


303 CIRCUITS

FIG. 56-1

This circuit uses a 15-kHz oscillator coil. When metal placed in the energy field is withdrawn, the oscillator voltage is rectified and compared to a reference. A drop in oscillator voltage therefore operates comparator IC2 and D4 (LED) extinguishes.

LOW-COST METAL DETECTOR FOR EXPERIMENTERS

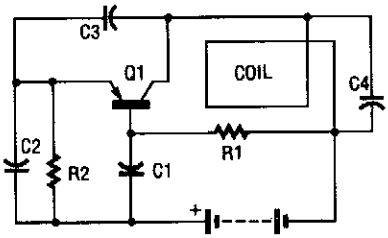


This circuit is on oscillator with L1 being a 4" diameter coil of 35 turns of #26 magnet wire. Metal in proximity to L1 will cause the oscillator to shift frequency. An AM transistor radio is used to detect the frequency shift.

POPULAR ELECTRONICS

FIG. 56-2

METAL LOCATOR



1991 PE HOBBYIST HANDBOOK *FIG. 56-3*

The metal locator uses a one-transistor oscillator and an AM radio to detect metal. Transistor Q1 is a pnp transistor that is connected to an oscillator. Resistor R1 provides the correct base bias and capacitors C3 and C4 and the search coil determine the frequency of oscillation.

Capacitors C3 and C4 are fixed in value, but the search coil is an inductor that varies in inductance (and thus varies the oscillator frequency) as metal is brought near it. The oscillator frequency is rich in harmonics and its output falls within the AM broadcast band. The metal detector works by combining its output with the local oscillator of the AM radio. The resulting net output of the radio is a low-frequency audio tone that changes—gets higher or lower—as metal is brought near or taken away from the search coil. Commercial metal detectors use two oscillators, so they don't require an AM radio. This metal locator provides an inexpensive alternative to an expensive commercial metal locator.

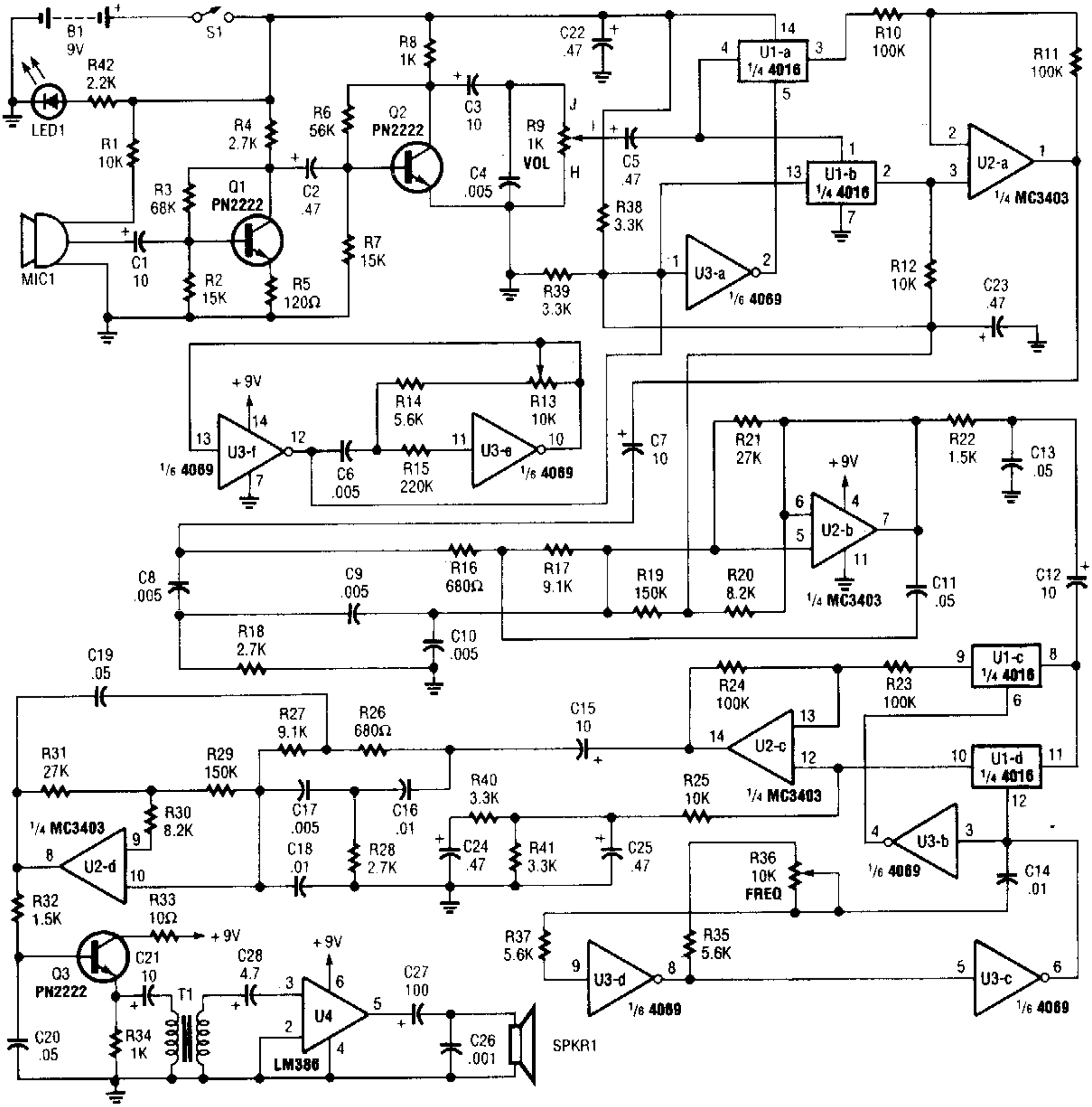
C1, C2	0.01-µF Capacitor (103)
C3, C4	0.001-µF Capacitor
Q1	2N3906 Transistor
R1	47-kΩ Resistor
R2	100-Ω Resistor

57

Miscellaneous Treasures

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

VOICE DISGUISER



POPULAR ELECTRONICS *FIG. 57-1*

A complete schematic diagram of the voice disguiser is shown. Microphone MIC1 picks up the voice signal and feeds it to an audio amplifier, consisting of Q1 and Q2, and a few support components. The amplifier has a low-pass gain response that limits the voice frequencies to 5 kHz or lower.

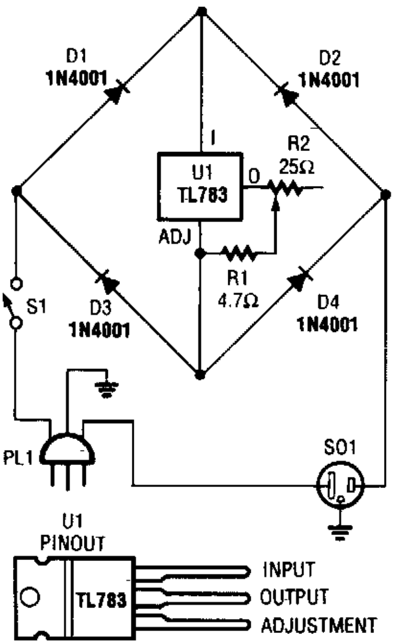
The voice signal is then fed to the input of the first balanced modulator, which is comprised of U1-a, U1-b, U2-a, and U3-a. The output of the first 4-kHz oscillator, built around U3-f and U3-e, is fed to the carrier input of the first modulator. The frequency of the first oscillator is controlled by the setting of potentiometer R13. The modulator output—a double-sideband suppressed-carrier signal centered on 4 kHz—is then filtered by the first 5-kHz low-pass filter, formed by U2-b, which eliminates the upper-sideband signals.

At this point, the voice frequency spectrum is inverted (e.g., the frequencies that were low now become high, and vice versa), making the voice signal completely unintelligible. The output of the first low-pass filter is fed to a second modulator formed by U1-c, U1-d, and U3-b, where it is frequency modulated with the output of the second carrier oscillator, comprised of U3-c and U3-d; the frequency of the second oscillator is controlled by potentiometer R36.

The output of the second modulator is filtered by the second low-pass filter, which consists of U2-d and few support components, and amplified by Q3. The voice output signal from Q3 is fed to U4 (an LM386 low-voltage, audio-power amplifier) through an impedance-matching transformer, T1. The output of U4 is then used to drive SPKR1 (an 8-Ω speaker).

In operation, if both carrier oscillators are set to the same frequency, the voice signal from the speaker will be an exact duplicate of the input signal from the microphone. However, if the frequency of the second oscillator is varied (via R36), the output voice signal also shifts in frequency. That makes the voice reproduced by the speaker sound higher- or lower-pitched than normal.

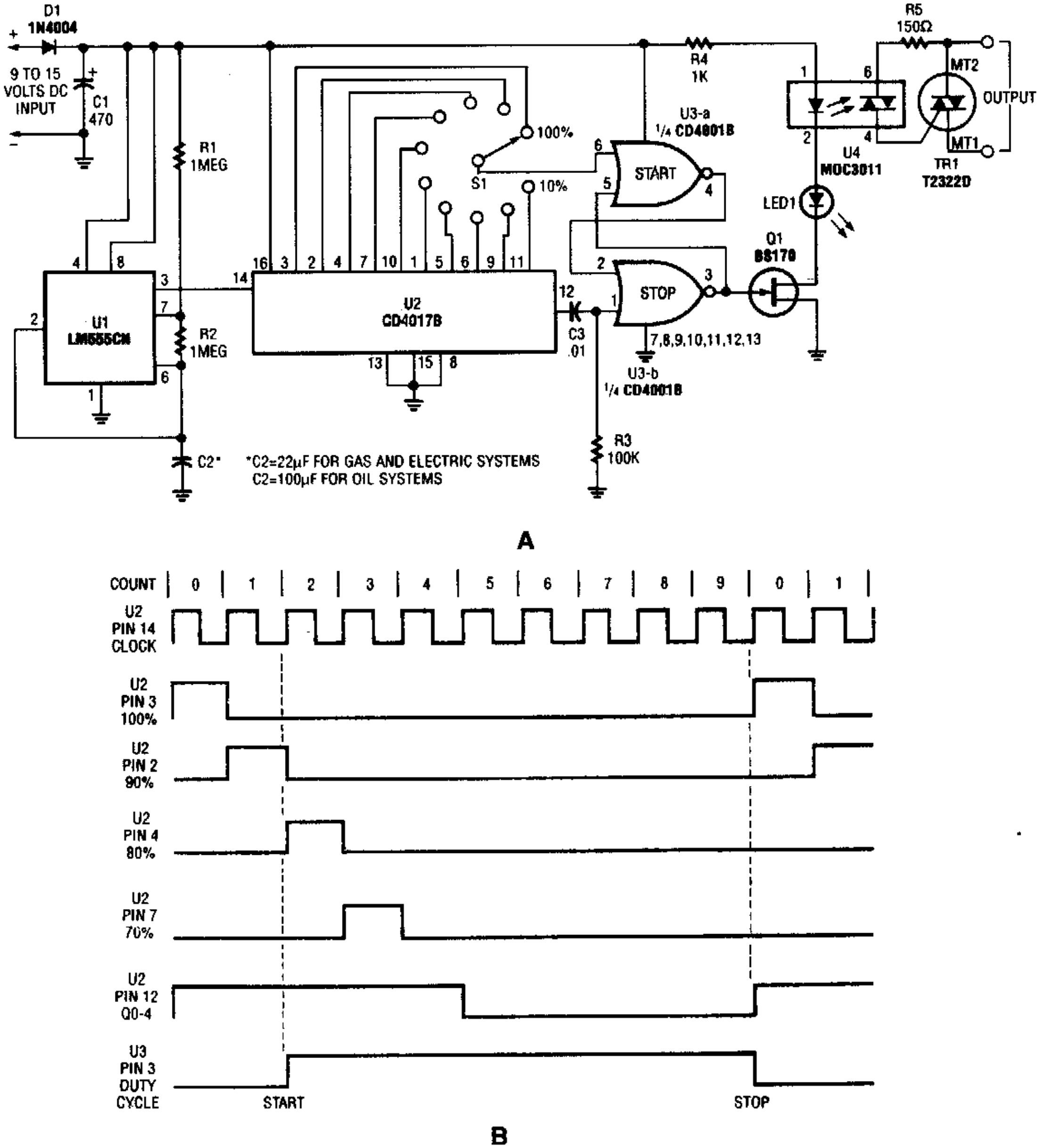
SOLDERING IRON CONTROL



A current control to temperature regulate a soldering iron uses a high-voltage integrated regulator, TL783 (U1). WIth the component values specified, the circuit should be used with a soldering iron of 25 W or less.

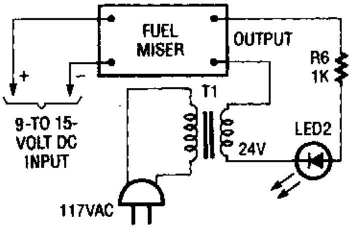
1992 ELECTRONICS HOBBYISTS HANDBOOK *FIG. 57-2*

FURNACE FUEL MISER

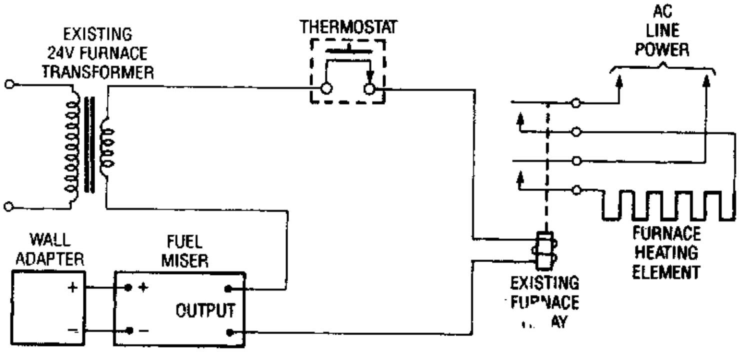


POPULAR ELECTRONICS ***FIG. 57-3***

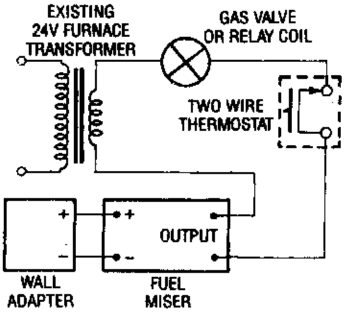
A timer (LM555CN) and decode counter is used to generate duty cycles from 10% to 100% to control the time a heating system can operate. V2 is a decode counter that can be switched from 10% to 100% duty cycle. V3A and B form a latch that drive A1, LED1, and V4. The triac TRI is used as an ac switch, in series with the thermostat that controls the heating system.



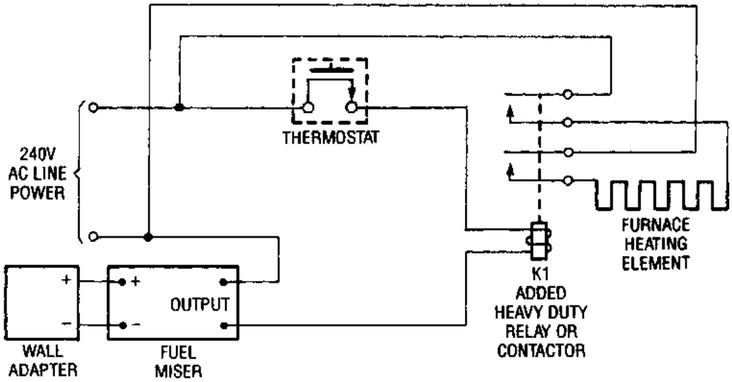
When the circuit is working properly, the output circuitry can be checked using a 24-volt step-down transformer, a 1k resistor, and an LED. Together those components simulate the load that the Fuel Miser sees during normal operation.



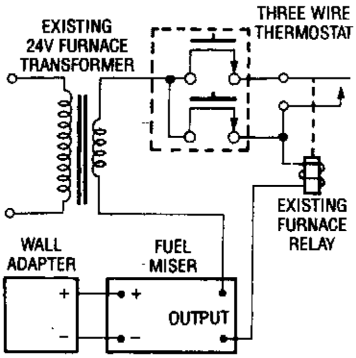
Electric-heating systems may or may not use a relay in the thermostat circuit. Those that do have a relay can be controlled by the Fuel Miser by wiring its output circuit in series with the relay coil connections as shown here.



This drawing shows the Fuel Miser connected in series with the thermostat of a two-wire gas furnace that's powered by a 24-volt transformer.

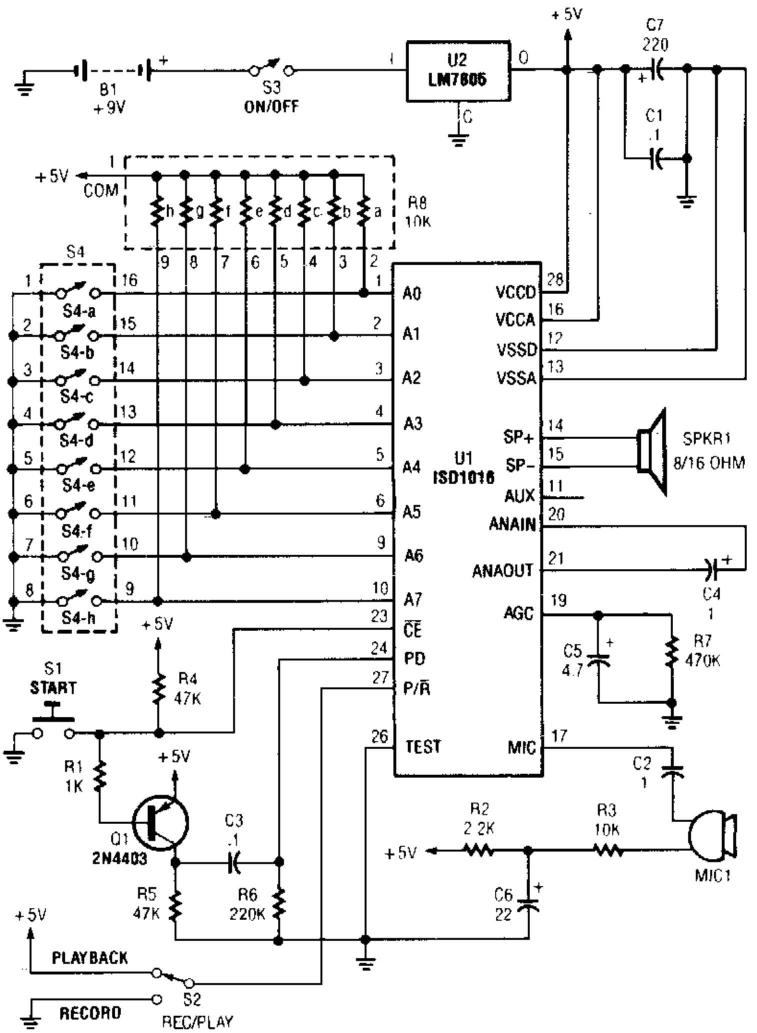


Electric-heating systems that do not contain a low-current thermostat (as in the previous installation), use a heavy-duty thermostat that directly feeds current to the heating element. For such systems, it will be necessary to install a heavy-duty relay (K1 in this example) to control the heavy heating-element current.



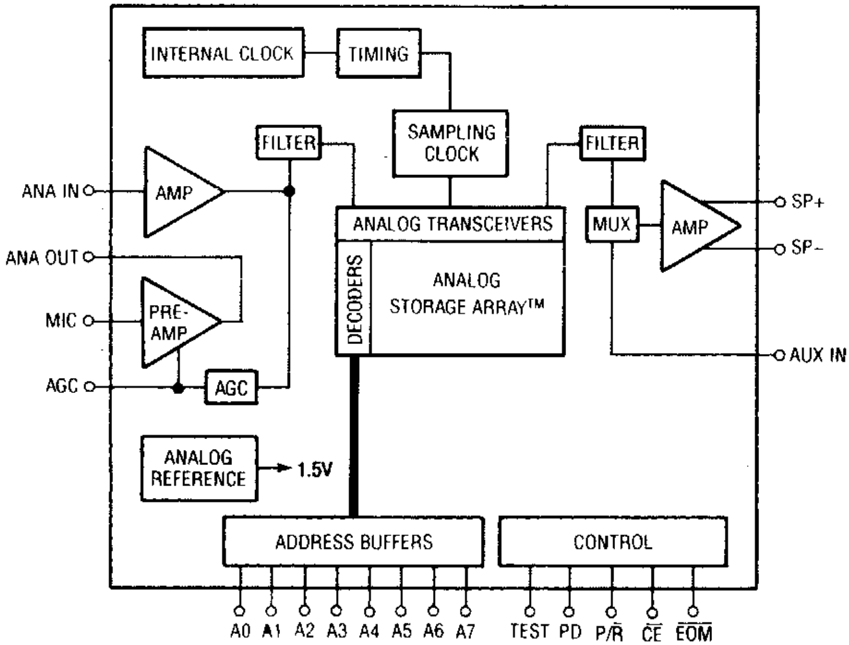
Some oil-fired systems use three-wire thermostats to control the operation of the burner motor and ignition system by activating a relay. This is a typical installation for such systems.

PERSONAL MESSAGE RECORDER



POPULAR ELECTRONICS

FIG. 57-4



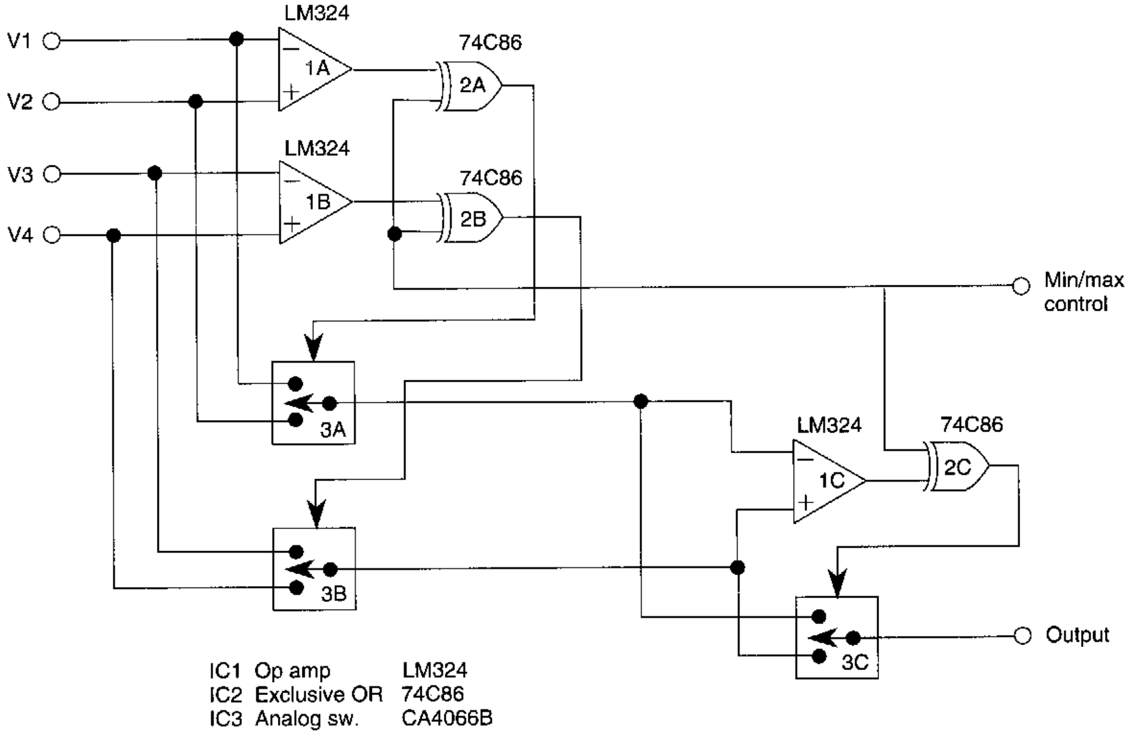
The personal message recorder is built around an ISD1016 CMOS voice messaging system, which does away with the cumbersome and expensive analog-to-digital and digital-to-analog conversion circuits.

A functional block diagram of the ISD1016 is shown. The ISD1016 contains all of the functions necessary for a complete message-storage system. The preamplifier stage accepts audio signals directly from an external microphone and routes the signals to the ANA OUT (analog out) terminal. An automatic-gain control (AGC) dynamically adjusts the preamplifier gain to extend the input signal range. Together, the preamp and AGC circuits provide a maximum gain of 24 dB. The internal clock samples the signal and, under the control of the address-decoding logic, writes the sampling to the analog-storage array. Eight external input lines allow the ISD1016's message space to be addressed in 160 equal segments, each with a 100-millisecond duration. When all address lines are held low, the storage array can hold a single, continuous, 16-second message.

However, there is a special addition to the POWER DOWN input (pin 24) of U1. If the internal memory becomes full during recording, an overflow condition is generated in order to trigger the next device. Once an overflow occurs, pin 24 must be taken high and then low again before a new playback of record operation can be started.

Transistor Q1, C3, R5, and R6 form a one-shot pulse generator that automatically clears any overflow condition each time that start switch (S1) is pressed. Switch S2 selects either the playback or the record mode. Switch S4—an 8-position (a–h) DIP switch—is included in the circuit to allow the circuit's record/playback time to be varied from 0 to 16 seconds. The maximum time available is when all 8 switch positions are closed (or set to the on position). Resistor network R8 (a–h) is included in the circuit to provide a pull-up function for the address lines, which thereby controls U1's record/playback time.

FOUR-INPUT MINIMUM/MAXIMUM SELECTOR



WILLIAM SHEETS

FIG. 57-5

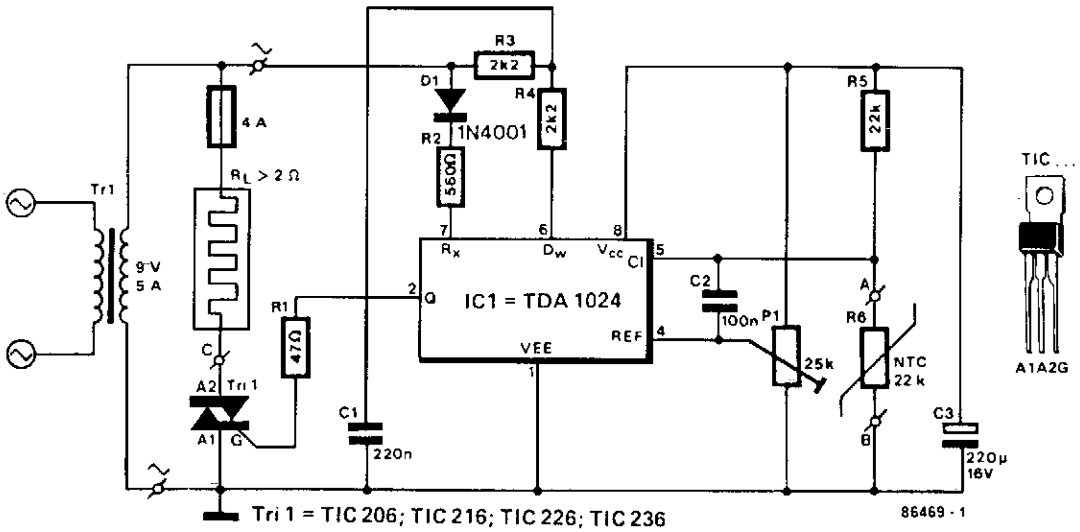
This circuit outputs the maximum (or the minimum) of the four input voltages V_1, V_2, V_3, and V_4. Each of these input voltages is in the range 0 to 5 V.

The output of the unit is the maximum of V_1, V_2, V_3, and V_4 if the control voltage input is 5 V (i.e., logical 1). The output is the minimum of V_1, V_2, V_3, and V_4 if the control input is zero.

By cascading N such units, one can select the maximum (or the minimum) of $3N + 1$ input voltages.

Thus if k is the number of input voltages, we need $[(k+1)/3]$ units.

SOIL HEATER FOR PLANTS

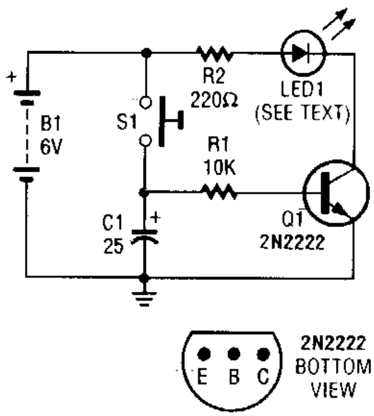


303 CIRCUITS

FIG. 57-6

A TDA1024 electronic thermostat senses soil temperature via thermistor R6. The circuit uses zero-crossing switching of the heater. The heater is made of elastic-coated steel wire. P1 is used to set the temperature. The heater should have 2 Ω or more resistance and operate from the 9-V transformer. About 40 W of heat is available.

KEY ILLUMINATOR



Used as a 10-second momentary illuminator, this circuit can be useful in other applications as well. Pressing S1 charges C1, which holds Q1 on and holds the LED lit for about 10 seconds.

POPULAR ELECTRONICS

FIG. 57-7

RADIO COMMERCIAL ZAPPER

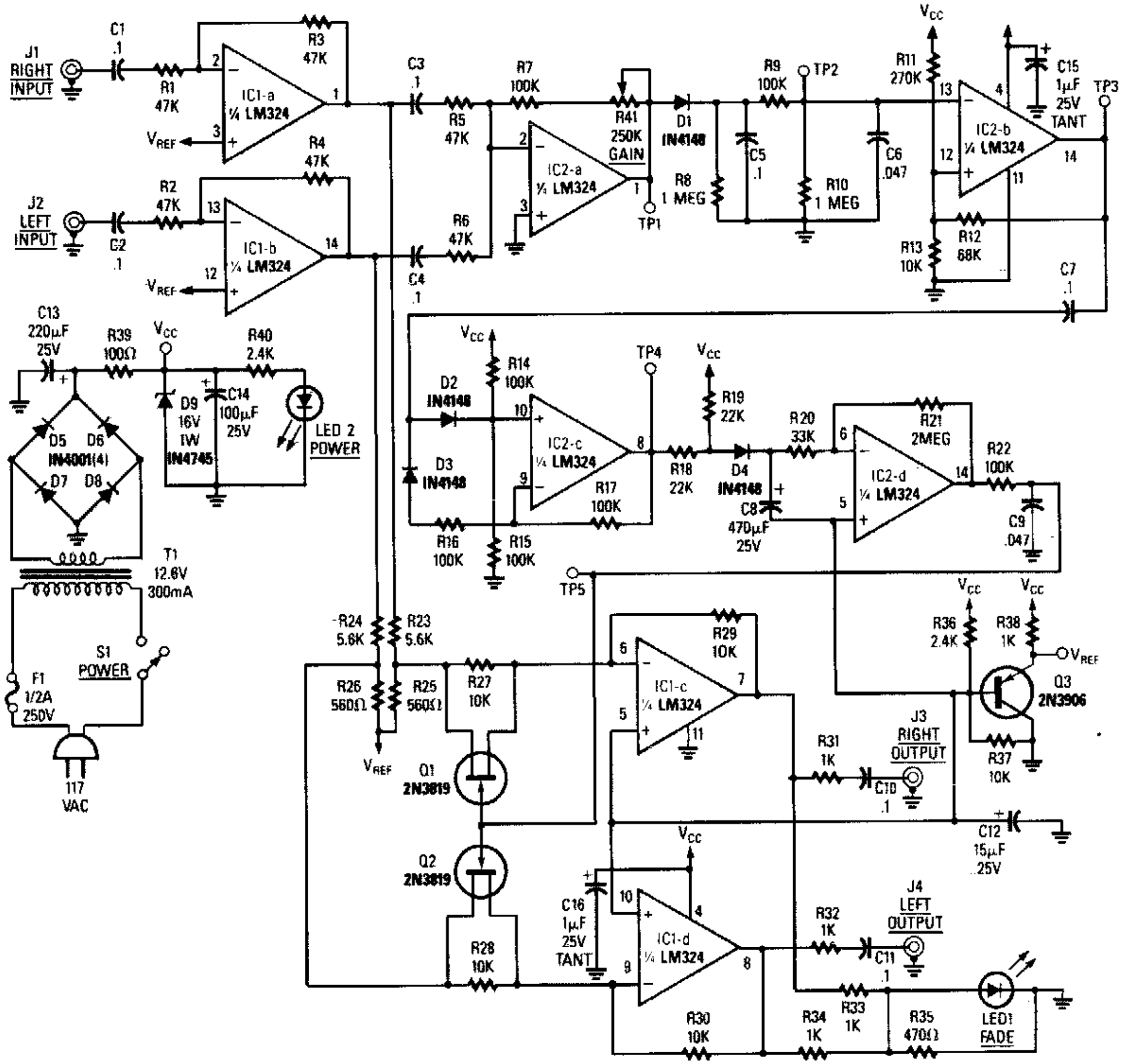
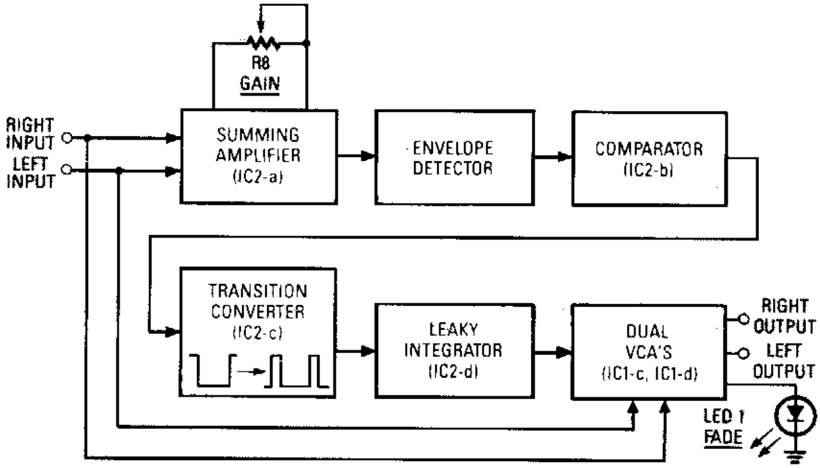


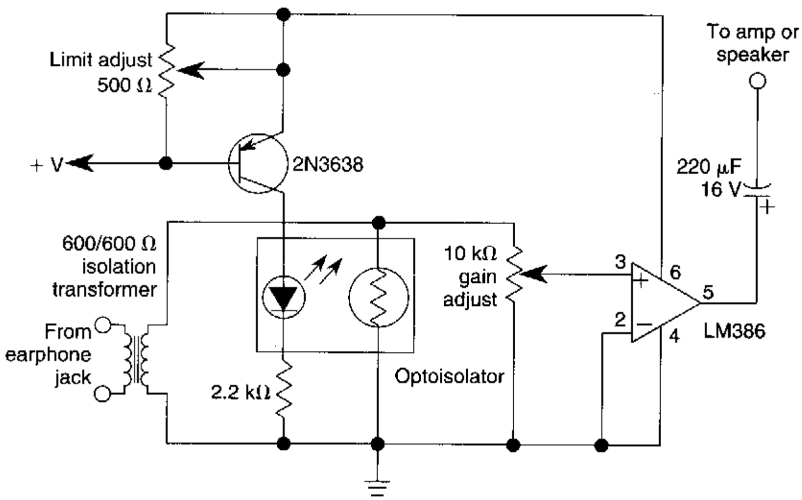
FIG. 57-8



BLOCK DIAGRAM OF THE COMMERCIAL KILLER: The envelope of the signal is used to vary the pulse rate from IC2-c. The pulses are integrated; the resulting signal controls the gains of a pair of VCA's.

The L&R inputs are summed, dated and drive a comparator. The comparator senses level and generates a transition when audio inputs go above or below preset thresholds. The number of these transitions (corresponding to rapid volume changes) are integrated and feed voltage controlled amplifiers. This device actually senses dynamic range.

AUDIO LIMITER



RADIO ELECTRONICS

FIG. 57-9

An optoisolator is used as an attenuator in this circuit. When the LM386 draws more current on audio signals, the 2N3638 turns on, which biases the optoisolator on, and reduces the volume.

ANALOG DE-GLITCH CIRCUIT

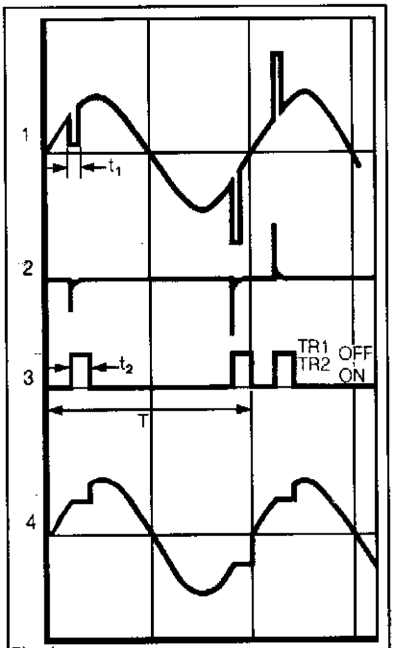


Fig. 1

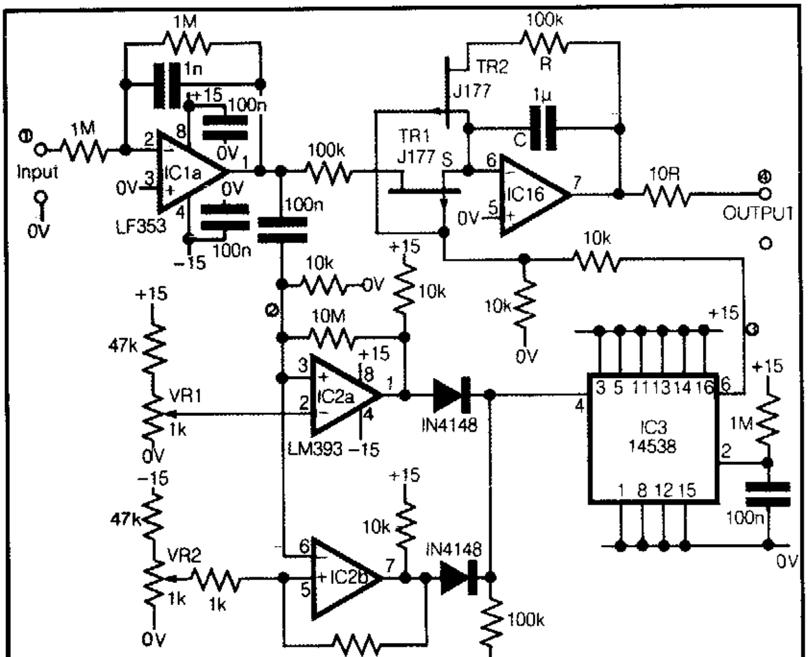


Fig. 2

FIG. 57-10

Low-frequency signals produced by transducers, measurement equipment, or data loggers often appear like the first waveform in the figure. The circuit shown operates as a tracking sample-hold, and the transients are replaced in the output by the stored value of the current signal at the instant of the transient.

The input signal is buffered and inverted by IC1a, and the differentiated result shown at 2 applied to the inputs of two comparators IC2-a and IC2-b. VR1 and VR2 set levels to prevent false or unnecessary operation. Either comparator output triggers the mono IC3 from positive or negative signal transients. When IC3 has not been triggered, TR1 and TR2 'p' channel JFETs are on, and IC1b operates as an integrator with a high leakage, and tracks the input signal. When the mono is triggered as at 3, TR1 and TR2 turn off and the previous signal value is held constant, as shown at 4. The resulting output waveform can then be easily filtered to remove the harmonics from the restoring step at the end of the mono period, if needed.

The criteria for successful operation are:

$t_2 > t_1$ (mono period longer than glitch)

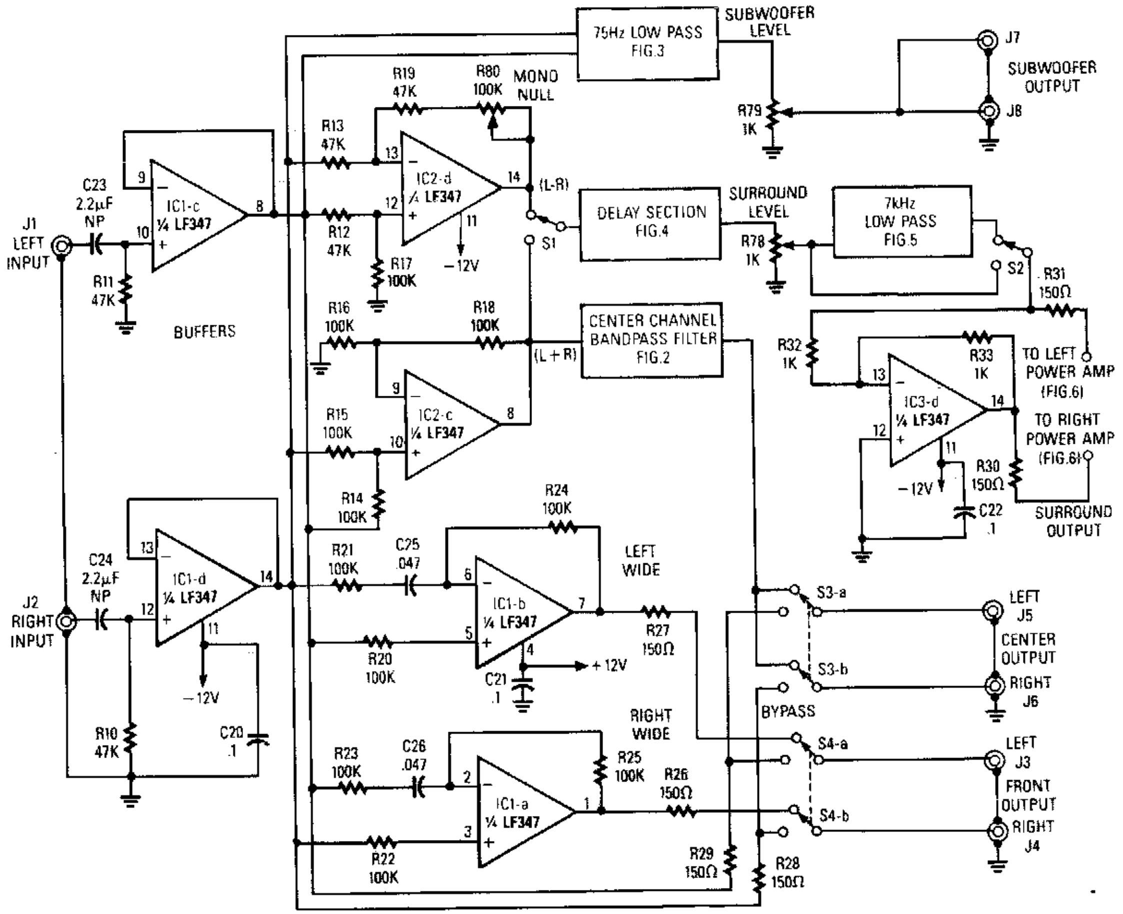
t_2/T small (to optimize output waveform)

Signal bandwidth $f_o = \dfrac{1}{2\pi CR}$

Signal phase $0 = \tan^{-1} 2\pi fCR$

The signal range is approximately ±5 V, depending on the transient amplitude and polarity. The mono period shown is 100 mS, but this can be optimized in practical applications. The shorter the mono period in relation to the signal waveform, the better the quality of the result.

ACOUSTIC FIELD GENERATOR



THE AFG IS MADE UP OF 10 relatively simple circuit elements.

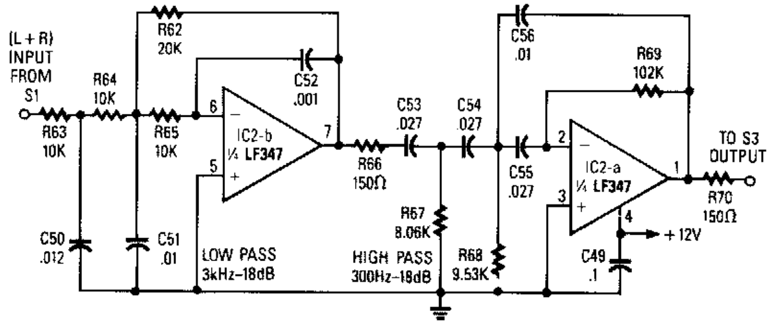
A

R-E EXPERIMENTER'S HANDBOOK ***FIG. 57-11***

Referring to the simplified schematic in A, the AFG is made up of 10 relatively simple circuit elements. IC1-c and IC1-d are configured as unity-gain noninverting buffer amplifiers.

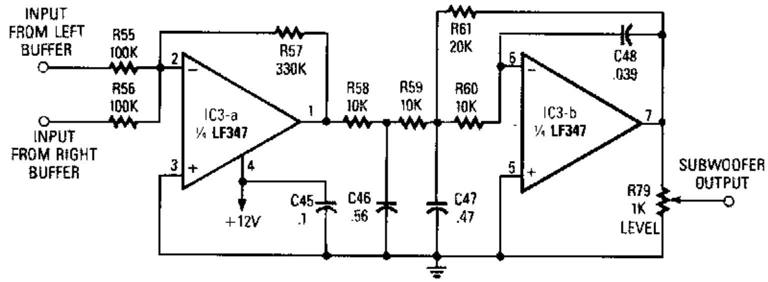
The summing (L+R) amplifier, IC2-c, combines equal amounts of the left and right signals, via R14 and R15, to develop a total composite signal. Left- and right-channel signals are applied equally through R13 and R12 to IC2-d, the difference (L–R) decoder. Any common to both channels is canceled by IC2-d, which exactly balances the inverting and noninverting gains of the amplifier for a perfect null.

The stereo width-enhancement circuit made up from IC1-a and IC-b works similarly to the (L–R) decoder, except that C25 and C26 have been added in the inverting inputs of each op amp. IC1-b develops the "left wide" signal because its inverting and noninverting inputs are connected to the left



THE CENTER-CHANNEL SPEECH FILTER is built by cascading a 3-kHz low-pass filter with a 300-Hz high-pass filter to form a band-pass filter.

B



AN ACTIVE CROSSOVER NETWORK for driving a high-power subwoofer system is made from IC3-a and IC3-b.

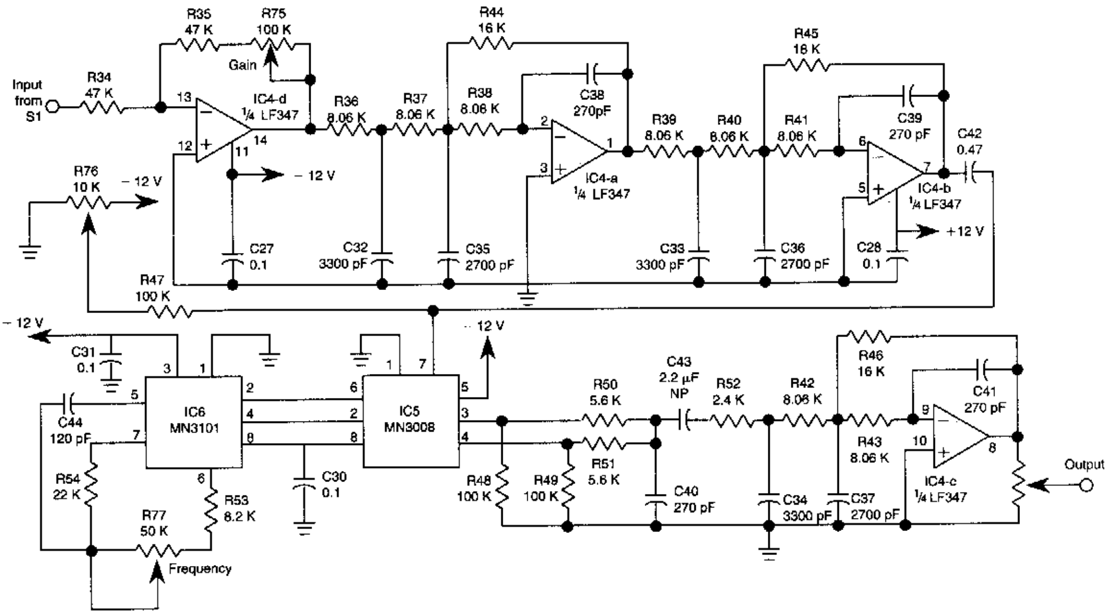
C

and right channels opposite that of IC1-a. The output of the width-enhancement circuit is routed to S4, which selects either the "wide" or the bypass signal for feeding the front-channel amplifier.

The center-channel dialogue filter is built by cascading a 3-kHz low-pass filter with a 3-Hz high-pass filter to form a band-pass filter. It has a sharp –18 dB/octave cutoff, a flat voltage and power frequency response, and minimum phase change within the passband.

In C, IC3-a and IC3-b form an active crossover network for driving a subwoofer. IC3-a sums signals from the left- and right-channel buffer amps, it inverts the summed signal 180 degrees, and provides a low driving impedance for the following filter stage. IC3-b and its associated RC network form a 75-Hz, 3rd-order low-pass filter. The filter inverts the signal another 180 degrees, so the signal that appears across R79 (which is the output-level control) is back in phase with the original input signal.

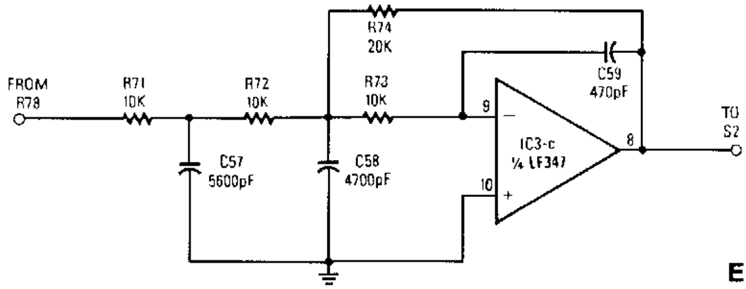
The delay section of the AFG, shown in D, is built around the MN3008 bucket brigade device (BBD), and the MN3101 two-phase variable-frequency clock generator. The amount of delay required in this system varies between approximately 5 to 35 milliseconds. The delay time of a BBD is equal to the number of stages divided by twice the clock frequency. Values were chosen for R53, R54, R77, and C44, to produce a clock frequency, adjustable via R77, which varies from about 30 kHz to 130 kHz.



D

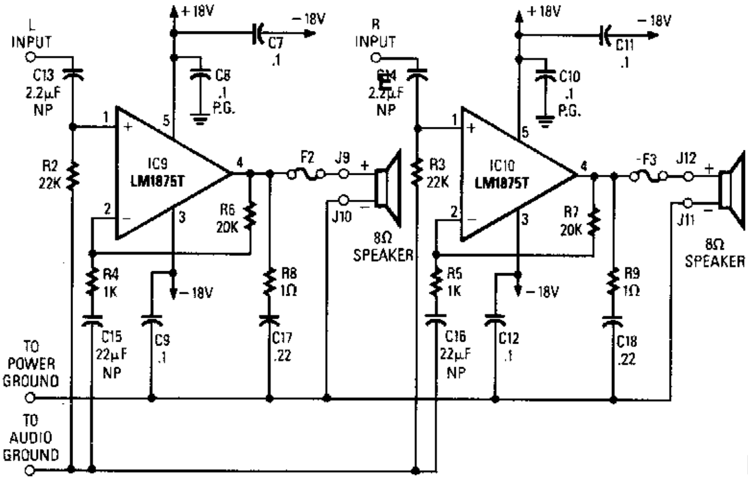
In A, S1 selects the signal to be delayed; either the difference signal (*L–R*) from IC2-d in the matrix mode or the sum signal (*L+R*) from IC2-c in the concert mode. The selected signal is fed from S1 to the delay section (D) where IC4-d is configured as an inverting amplifier; R75 adjusts the gain between unity and X3. Integrated circuits IC4-a and IC4-b, along with their assoicated RC networks, are identical 3rd-order 15-kHz low-pass filters. Cascading two filters produces a very sharp cut off (–36 dB per octave). Potentiometer R76 adjusts the bias voltage required by the BBD to exactly one half the supply voltage, as required.

The power supply of the AFG, shown in G, is of conventional design. A 25-V center-tapped transformer, along with diodes D1 and D2, produces about ±18-V unregulated dc. Two 2200-µF filter capacitors provide ample energy storage to meet the high-current demands of the audio output amplifier ICs during high output peaks.



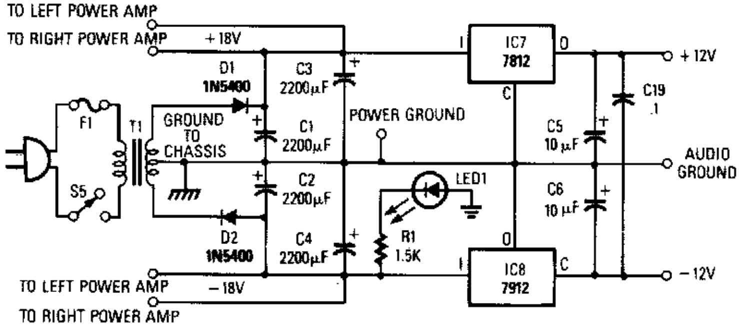
E

A 3rd-ORDER 7-kHz LOW-PASS FILTER is made from IC3-c and its associated RC network.



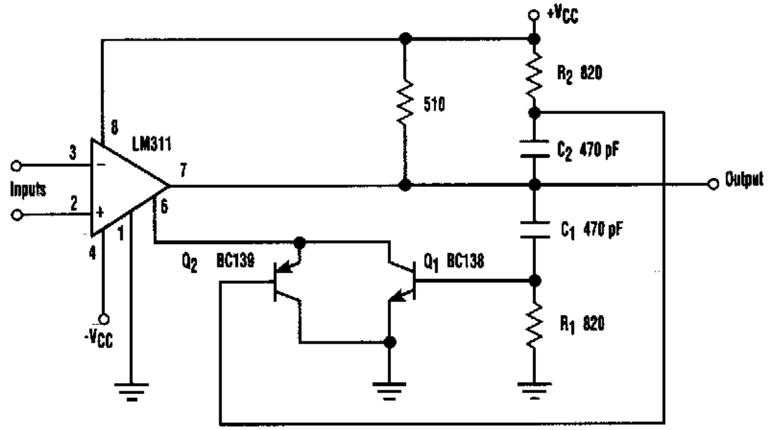
F

THE SURROUND CHANNEL POWER AMPLIFIERS are designed around a pair of LM1875 monolithic power-amplifier IC's.



THE POWER SUPPLY produces about ± 18-volts unregulated DC. G

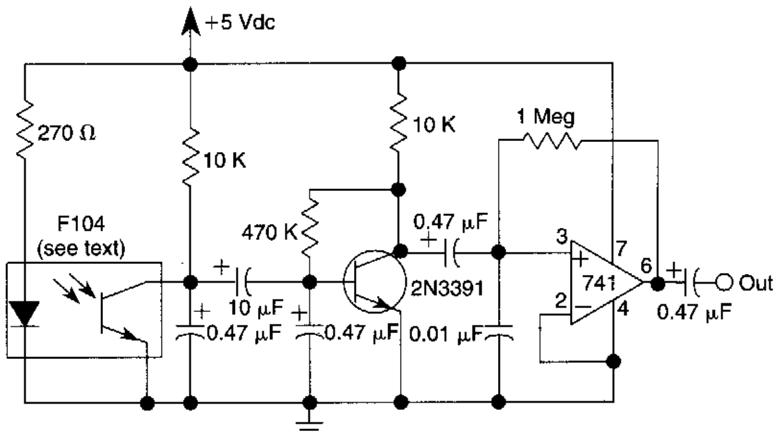
SUPPRESS JITTER WITH HYSTERESIS



ELECTRONIC DESIGN *FIG. 57-12*

When the comparator's output changes its state from low to high, the rising edge of the output pulse, differentiated by the C1/R1 chain, opens Q1. This blocks comparator M via its strobing input and sustains its output in the H state for a period of time, defined by the time constant R_1C_1. After C1 is charged by the current flowing through R1, Q1 is shut off and the comparator is released. When the comparator's output state changes from high to low, a similar process, involving elements R2, C2, and Q2, occurs. In many applications, the output transition in only one direction is of vital importance, and the elements, which provide temporal hysteresis for the opposite direction transition, can be omitted.

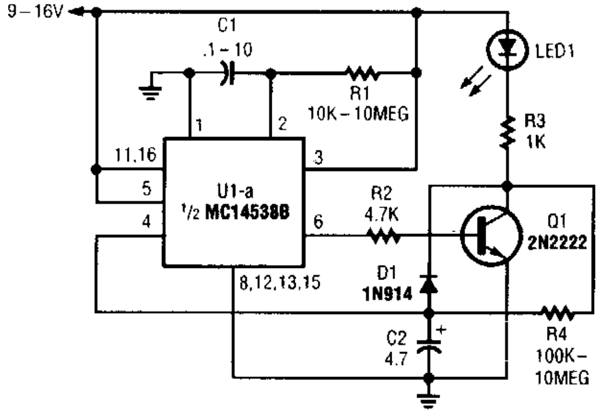
HEARTBEAT MONITOR



ELECTRONICS NOW *FIG. 57-13*

An IR photodiode, which senses IR skin reflectivity as a result of increased blood volume during the periods that the heart forcibly contracts, is used to pick up a signal that is correlated with the heartbeat. A transistor and op amp raise this to a level suitable to trigger logic circuitry or to be displayed on a scope.

SELF-RETRIGGERING TIMED-ON GENERATOR

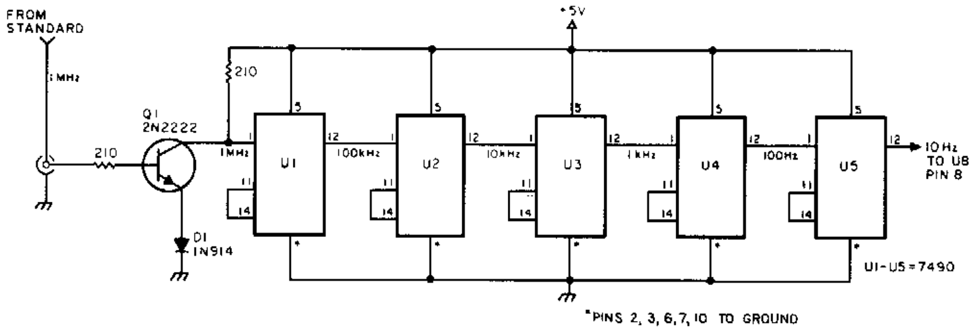


POPULAR ELECTRONICS **FIG. 57-14**

When power is first applied to the circuit, C2 begins to charge via LED1, R3, and R4. When the voltage across C2 reaches U1's input trigger level, the output of U1 at pin 6 goes positive for a period that is determined by the values of C_1 and R_1. That turns Q1 on, discharging C2 through D1 and Q1.

At the end of the set period, the output of U1 at pin 6 goes low, turning Q1 off and allowing the current to begin flowing through LED1, R3, and R4 to gain charge C2, causing the cycle to repeat. The repeat time is determined by the values of R_3, R_4, and C_2. The previous formula won't be as accurate for this circuit, but it will at least get you close enough for the capacitor value; then R_4 can be fine-tuned to obtain the desired timing period.

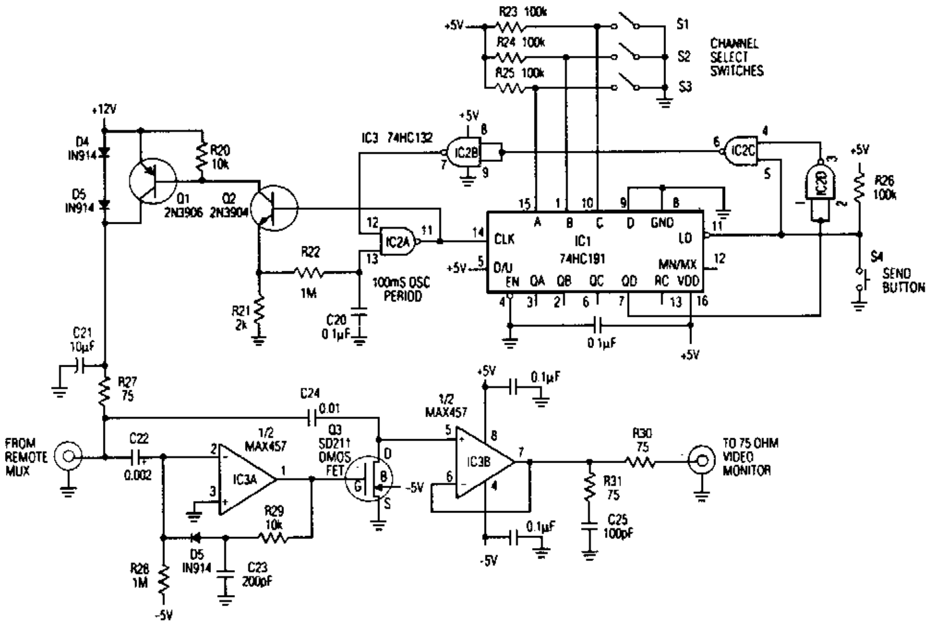
FREQUENCY DIVIDER FOR MEASUREMENTS



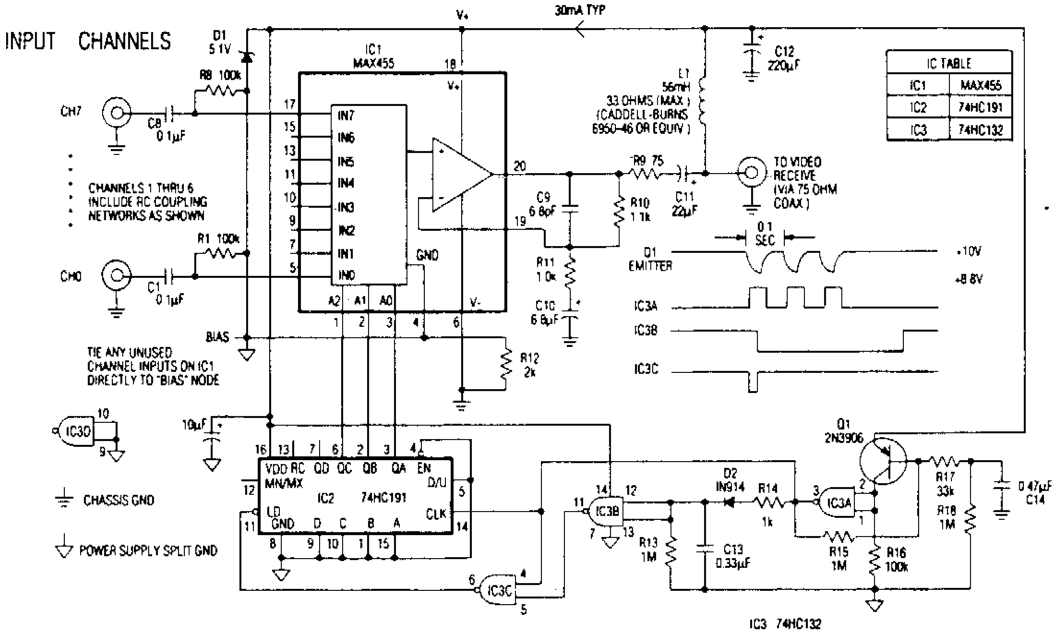
73 AMATEUR RADIO TODAY **FIG. 57-15**

This circuit is meant to be driven by a 1-MHz standard signal of a few volts amplitude. U1 through U5 are 7490 decade counter/divider and produce a division ratio of 100,000:1. Successive divisions of 10 can be tapped off, if desired, between stages. One or more stages can be added for still lower frequencies.

VIDEO, POWER, AND CHANNEL-SELECT SIGNAL CARRIER



A



B

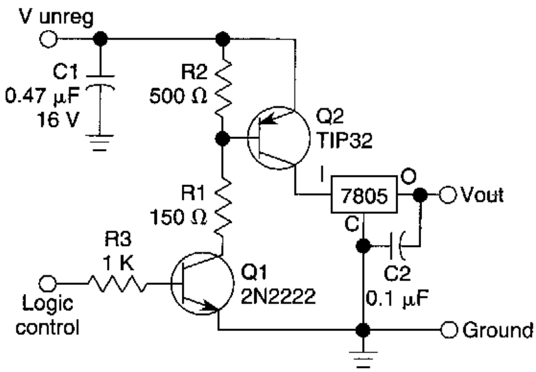
FIG. 57-16

In the video system of Figs. A and B, a single coaxial cable carries power to the remote location, selects one of eight video channels, and returns the selected signal. The system can choose one of several remote surveillance-camera signals, for example, and display the picture on a monitor near the interface box.

The heart of the multiplexer box (A) is a combination 8-channel multiplexer and amplifier (IC1). C11 couples the multiplexer's baseband video output to the coax, and L1 decouples the video from dc power arriving on the same line. This power—approximately 30 mA at 10 V—supplies all circuitry in the multiplexer box.

In interface box (B), a desired channel is encoded by three bits, set either by switches as shown or by an applied digital input. Momentary depression of the send button triggers downconverter IC1 and gated oscillator IC2A to initiate a channel-selection burst.

7805 TURN-ON CIRCUIT

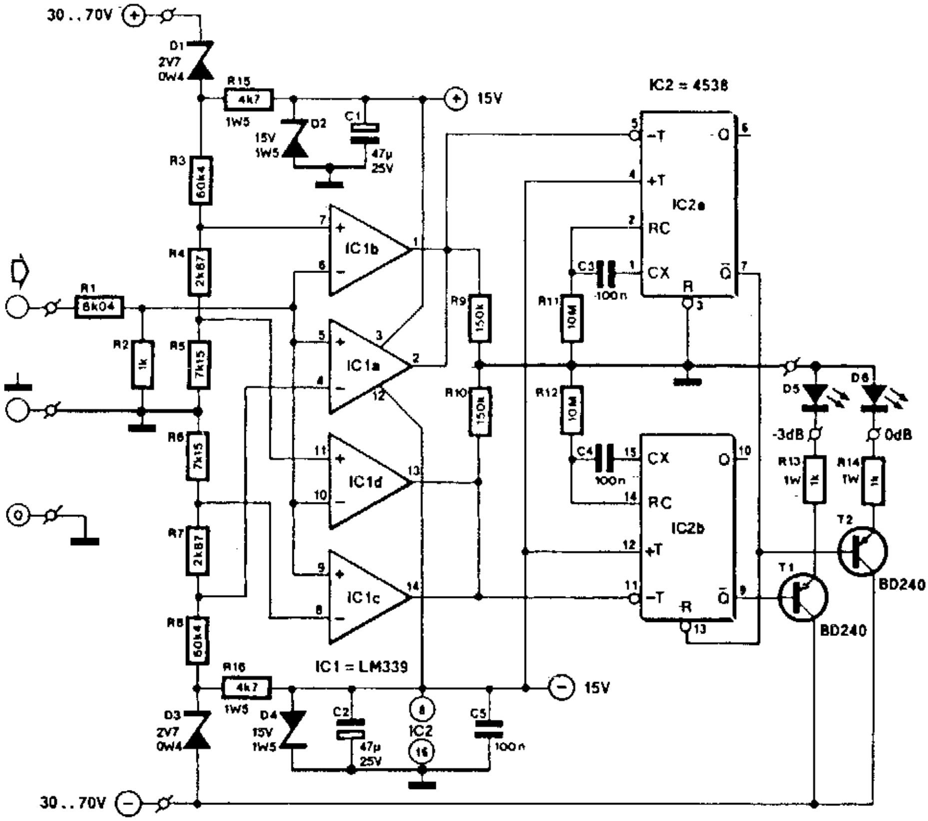


A logic level can control a 7805 regulator with this circuit. Q2 is a series switching transistor controlled by Q1. Q1 is turned on by a logic voltage to its base.

RADIO ELECTRONICS

FIG. 57-17

AF DRIVE INDICATOR

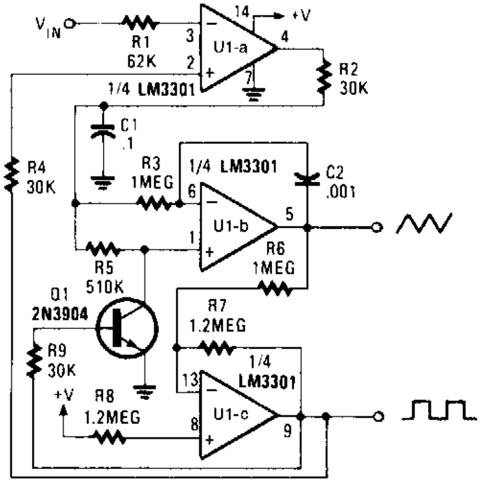


ELEKTOR ELECTRONICS

FIG. 57-18

This circuit was used with an audio power amplifier to detect the point at which output is –3 dB from maximum, indicated by LED D5, and at clipping, shown by LED D6. The indicator can be used with any amplifier operating from a ±30 to ±70 V symmetrical supply.

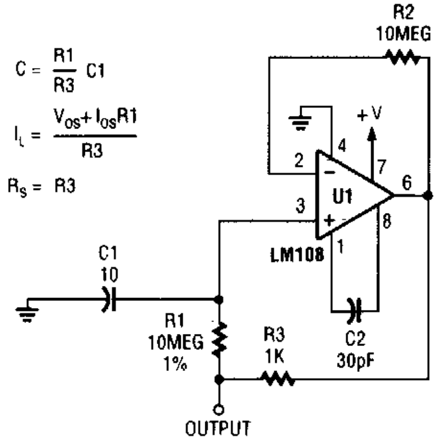
PHASE-LOCKED LOOP



POPULAR ELECTRONICS ***FIG. 57-19***

The PLL will lock onto an input signal. Both triangle- and square-wave outputs are available. A quad op amp can be used in this circuit, which should be useful in the audio and LF radio region.

CAPACITANCE MULTIPLIER



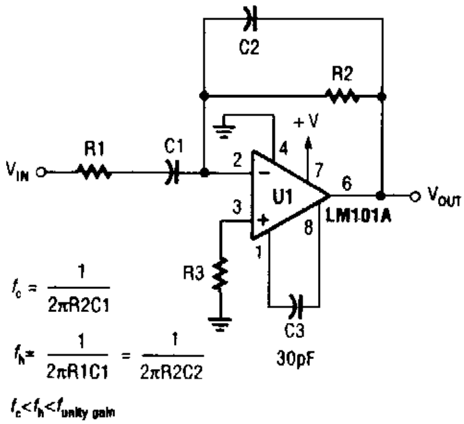
$$C = \frac{R1}{R3} C1$$

$$I_L = \frac{V_{OS} + I_{OS}R1}{R3}$$

$$R_S = R3$$

POPULAR ELECTRONICS ***FIG. 57-20***

PRACTICAL DIFFERENTIATOR



$$f_c = \frac{1}{2\pi R2C1}$$

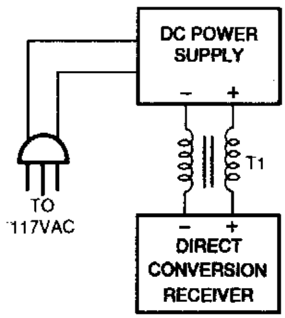
$$f_h = \frac{1}{2\pi R1C1} = \frac{1}{2\pi R2C2}$$

$$f_c < f_h < f_{\text{unity gain}}$$

POPULAR ELECTRONICS ***FIG. 57-21***

A differentiator has a high-pass characteristic. Components are chosen by using the design equations.

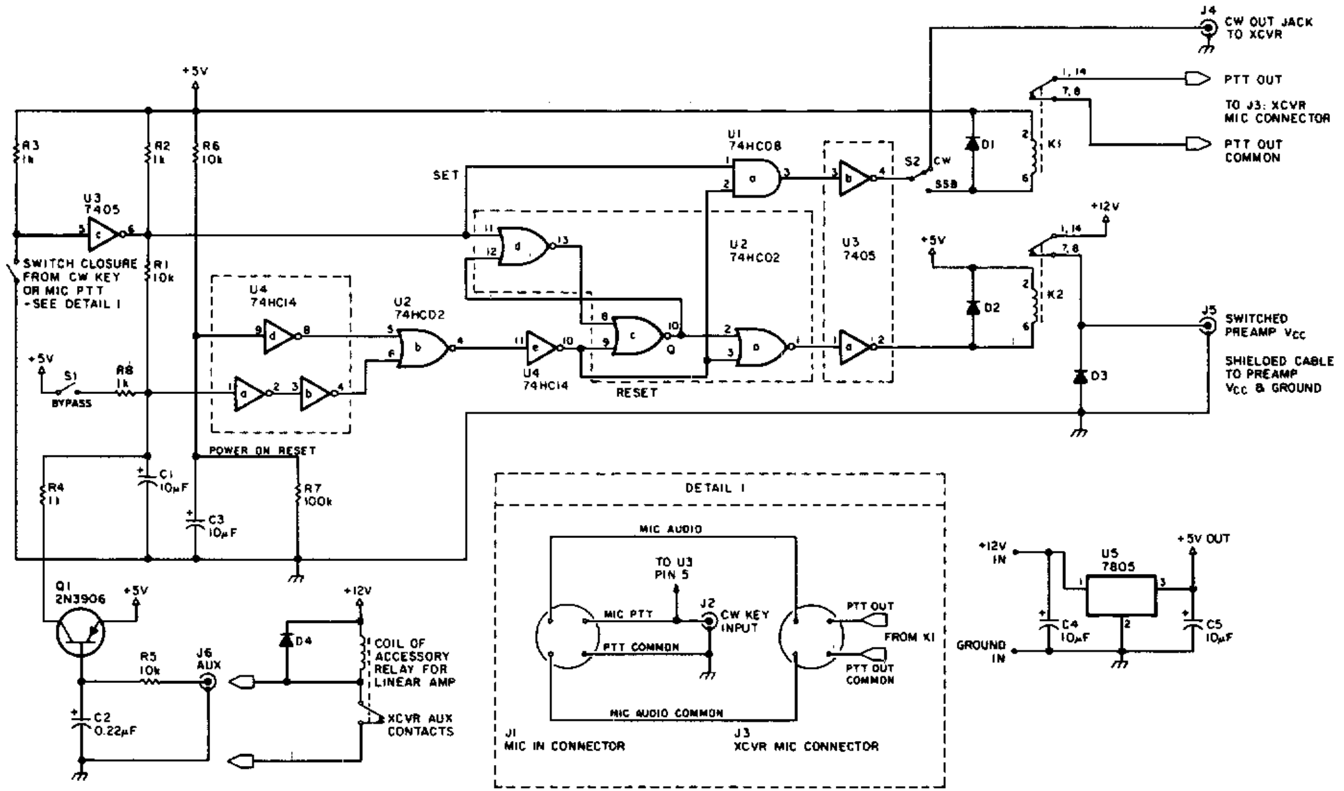
HUM REDUCER FOR DIRECT-CONVERSION RECEIVERS



POPULAR ELECTRONICS ***FIG. 57-22***

One cure for ac power line hum and ripple (caused by leakage current) is to use a well-regulated and filtered 9- to 18-Vdc power supply with a balancing choke (T1 in this illustration) between the power supply and the DCR.

PREAMP TRANSMIT-RECEIVE SEQUENCER

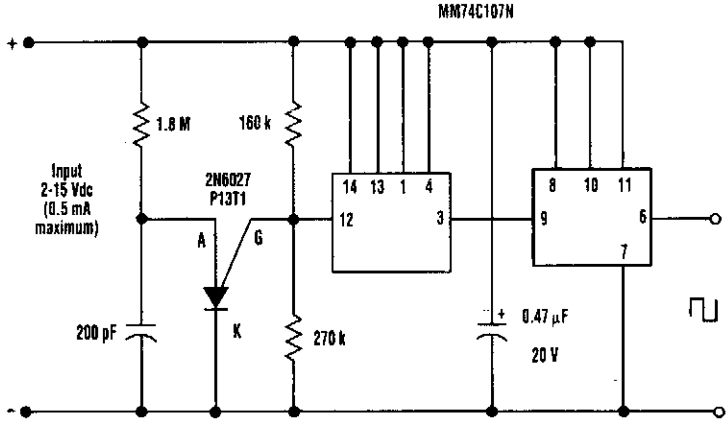


73 AMATEUR RADIO TODAY

FIG. 57-23

This circuit is useful in amateur radio VHF and UHF work where a mast-mounted antenna preamp is used for receiving. The kit controls T-R switching and change-over relay sequencing so that high RF levels are prevented from accidentally being applied to the preamplifier during switching intervals.

dc OUTPUT CHOPPER

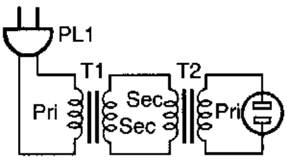


ELECTRONIC DESIGN **FIG. 57-24**

Any dc voltage source in the 2- to 15-V range can be chopped into a unipolar square wave that has a peak amplitude nearly equal to the dc source voltage with circuit (lightly loaded CMOS will swing within a few millivolts of each rail at low frequencies). Depending on the actual voltage of the supply, the programmable-unijunction-transistor (PUT) relaxation oscillator produces 2000-Hz trigger pulses. These pulses operate the cascaded 74C107 flip-flop, producing a square wave.

ac ISOLATION TRANSFORMERS USE INEXPENSIVE 12-V TRANSFORMERS

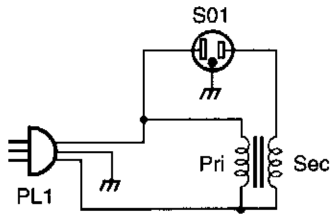
"Safety first" is a good motto to follow when you play with electricity. You can follow that adage more closely with this homebrew isolation transformer.



POPULAR ELECTRONICS **FIG. 57-25**

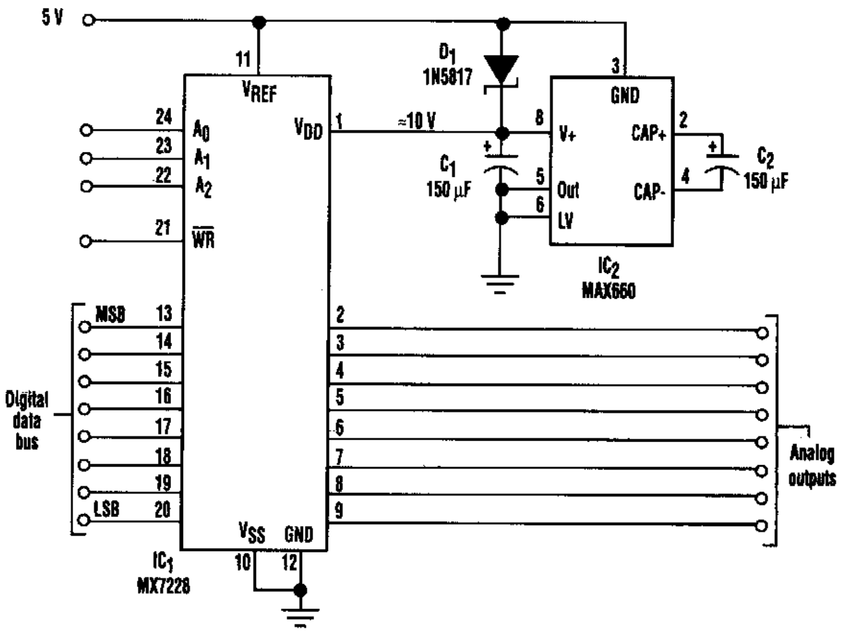
ac LINE VOLTAGE BOOSTER

When incoming ac power drops, you can bring the voltage back up with this booster circuit. It adds the transformer's secondary voltage to the ac line voltage.



POPULAR ELECTRONICS **FIG. 57-26**

OCTAL D/A CONVERTER

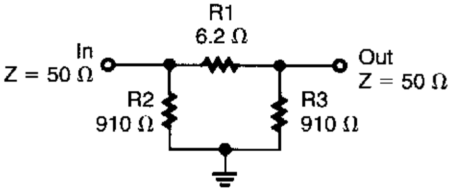


ELECTRONIC DESIGN

FIG. 57-27

This octal digital-to-analog converter operates on 5 V and provides eight output voltages, each digitally adjustable from supply rail to supply rail (0 to 5 V). Each output's resolution is 20 mV/LSB. The DAC chip (IC1) requires 3.5 V of "headroom" between its V_{DD} and reference voltages. However, a voltage-doubler charge pump (IC2) removes this limitation by generating an approximate 10-V supply for V_{DD}. All of the converter references are connected to the 5-V supply. IC2 doubles the 5-V input to an unregulated 10-V output that has an output impedance of less than 10 Ω. It can deliver 100 mA, which enables the eight DACs to issue their maximum output currents simultaneously (8 × 5 mA = 40 mA).

1-dB PAD

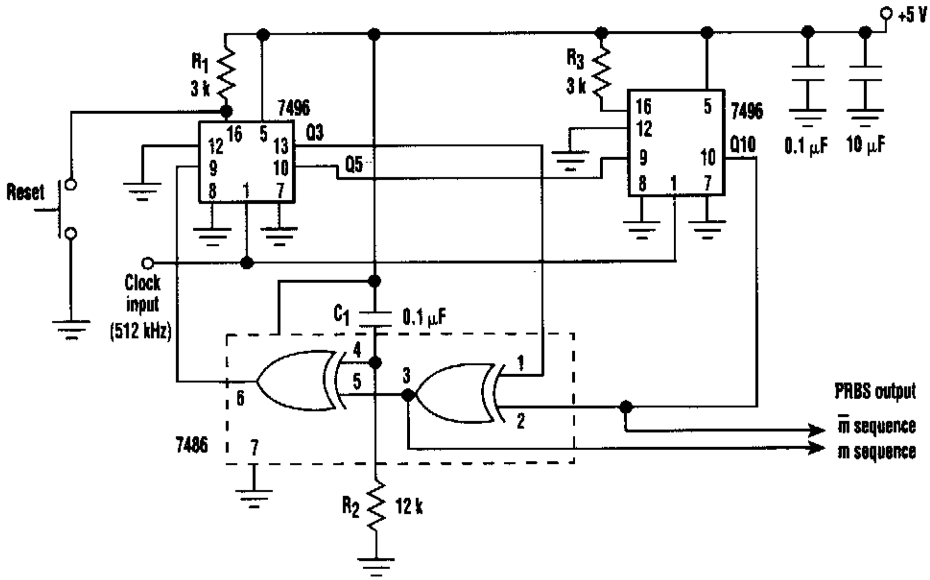


The 1-dB pad is useful as a termination in RF work to limit possible mismatch range between system blocks, etc.

POPULAR ELECTRONICS

FIG. 57-28

PSEUDO-RANDOM BIT SEQUENCE GENERATOR



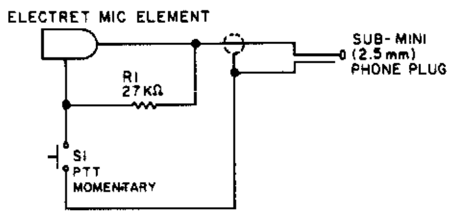
ELECTRONIC DESIGN

FIG. 57-29

In this circuit, an additional exclusive-OR gate is connected after the modulo-2 feedback, with C1 and R2 applying the supply turn-on ramp into the feedback loop. This provides sufficient transient signal so that the PRBS generator can self-start a power-up. A shift-register length n of 10 is shown with feedback at stages 3 and 10, providing true and inverted maximal length sequence outputs.

This technique applies an input directly to the feedback loop. Therefore, it's considered more reliable than applying an RC configuration to the shift-register reset input to create a random turn-on state.

SIMPLE EXTERNAL MICROPHONE CIRCUIT FOR TRANSCEIVERS

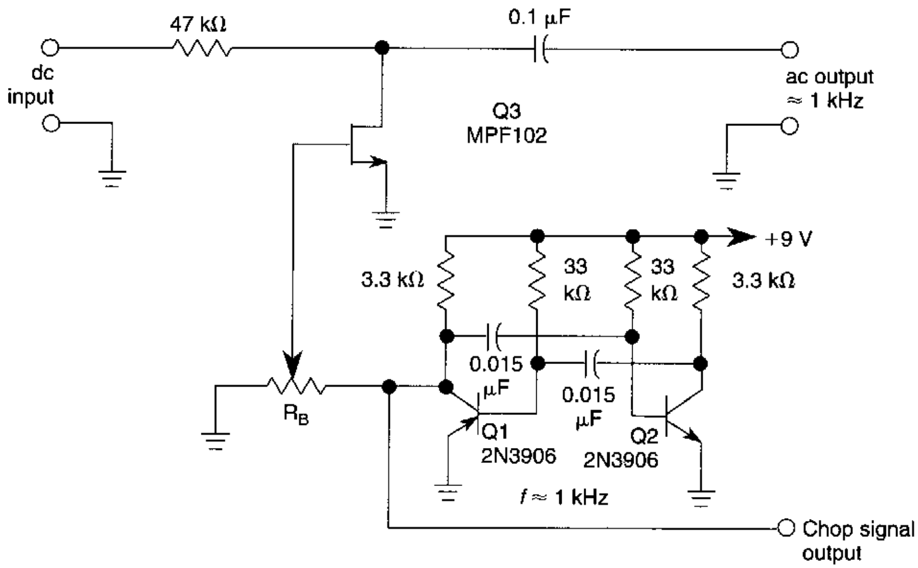


Used originally for an Icom ICZAT handie talkie, this circuit might prove useful in other applications.

73 AMATEUR RADIO

FIG. 57-30

JFET CHOPPER CIRCUIT

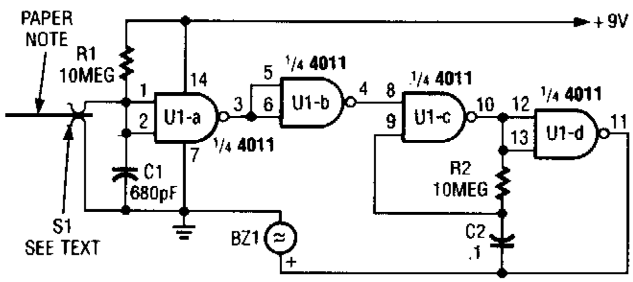


WILLIAM SHEETS

FIG. 57-31

A JFET (MPF102) is used to chop a dc signal for amplification in an ac coupled amplifier. Q3 is the chopper element and Q1-Q2 forms the multivibrator to derive a chopping signal. R_B sets the bias on the FET to keep the drive to MPF102 as low as possible.

AUDIO MEMO ALERT

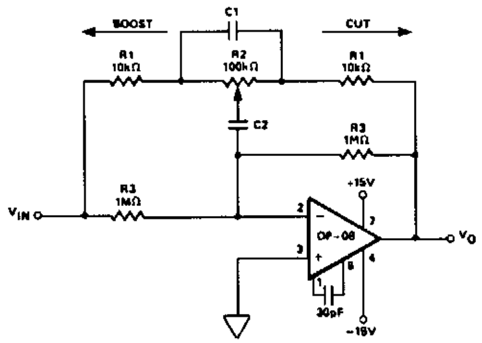


POPULAR ELECTRONICS

FIG. 57-32

This device prevents paper notes and memos from being overlooked. A paper note placed between two fingers made of a conducting material (metal or conductive plastic) breaks the circuit, allowing pair 1 of U1-a to go high. This causes U1-c & U1-d to act as an oscillator, pulsing piezo buzzer BZ1.

OCTAVE EQUALIZER



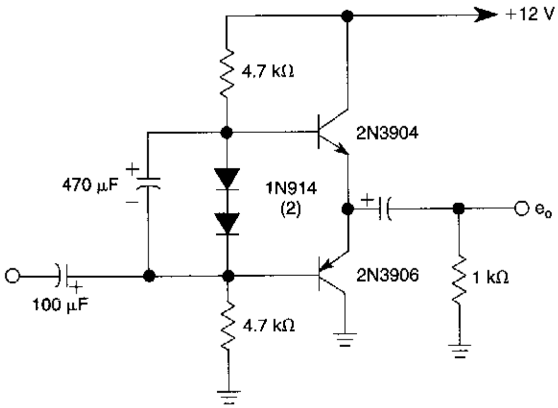
f_o (Hz)	C_1	C_2
32	0.18μF	0.018μF
64	0.1μF	0.01μF
125	0.047μF	0.0047μF
250	0.022μF	0.0022μF
500	0.012μF	0.0012μF
1k	0.0056μF	560pF
2k	0.0027μF	270pF
4k	0.0015μF	150pF
8k	680pF	68pF
16k	360pF	36pF

PRECISION MONOLITHICS INC.

FIG. 57-33

This circuit is one section of an octave equalizer used in audio systems. The table shows the values of C1 and C2 that are needed to achieve the given center frequencies. This circuit is capable of 12 dB boost or cut, as determined by the position of R2. Because of the low input bias current of the OP-08, the resistors could be scaled up by a factor of 10, and thereby reduce the values of C1 and C2 at the low-frequency end. In addition, 10 sections will only draw a combined supply current of 6 mA maximum.

COMPLEMENTARY OR BILATERAL ac EMITTER-FOLLOWER CIRCUIT

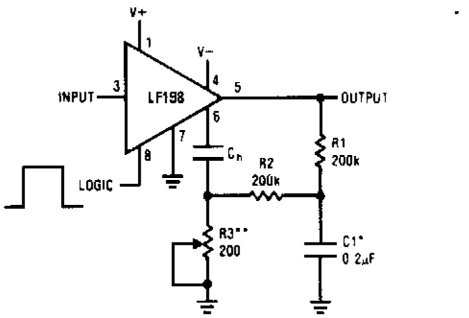


WILLIAM SHEETS

FIG. 57-34

This noninverting circuit uses a pair of complementary npn (2N3904) and pnp (2N3906) transistors.

CAPACITOR HYSTERESIS COMPENSATOR



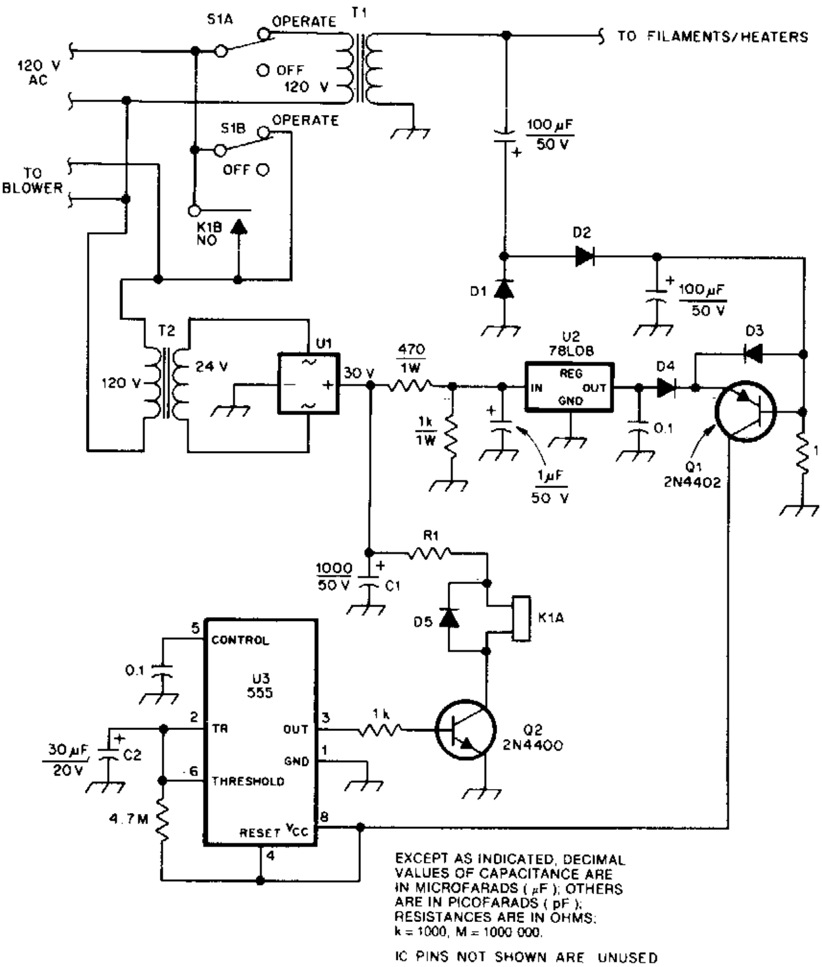
*SELECT FOR TIME CONSTANT C1 = $\frac{\tau}{100k}$

**ADJUST FOR AMPLITUDE

LINEAR DATABOOK

FIG. 57-35

AMPLIFIER COOL-DOWN CIRCUIT I

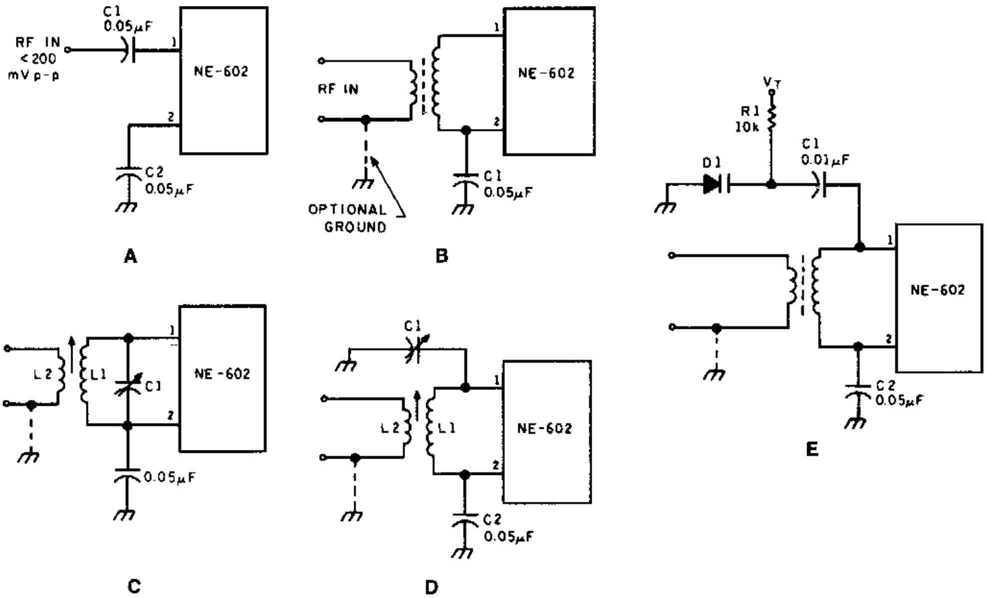


QST

FIG. 57-36

This cool-down relay circuit uses an IC timer to drive a relay, which keeps the blower on for a time delay from timer U3. The value of C_2 can be changed to lengthen or shorten the time, as needed.

NE602 INPUT CIRCUITS

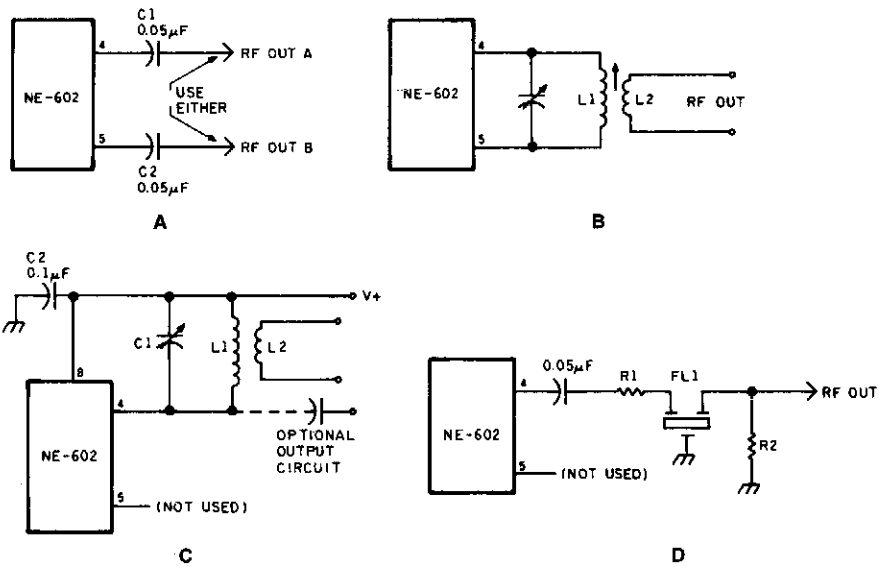


73 AMATEUR RADIO TODAY

FIG. 57-37

Input circuits for the NE-602.

NE602 OUTPUT CIRCUITS

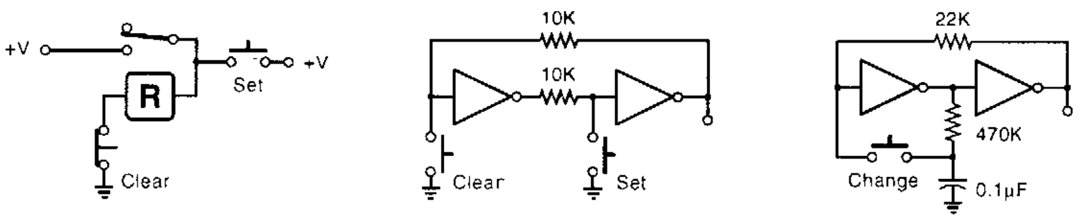


73 AMATEUR RADIO TODAY

FIG. 57-38

Output circuits for the NE-602.

BASIC LATCH CIRCUITS



(A) Relay converted to latch.

(B) Inverter pair used as latch.

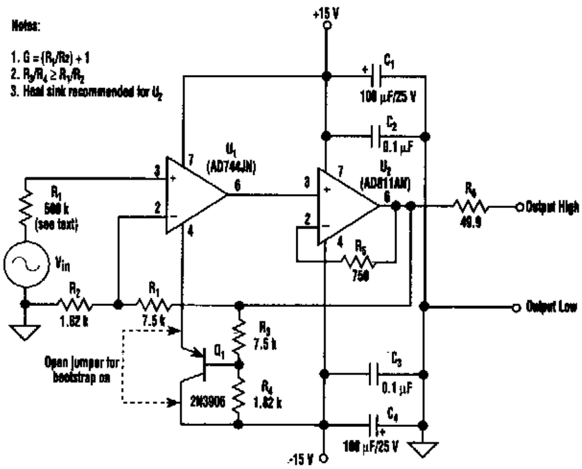
(C) Alternate action pushbutton.

ELECTRONICS NOW

FIG. 57-39

Some simple latches and alternate action circuits.

BOOTSTRAP CIRCUIT

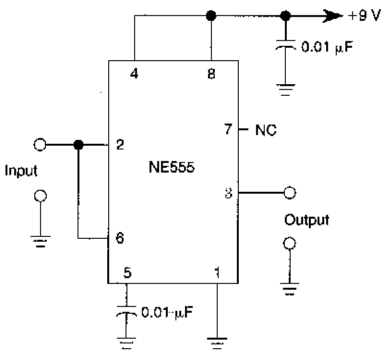


Bootstrapping the substrate of a JFET amplifier reduces the distortion caused by the nonlinearity of the JFET input capacitance. In the figure, a second feedback divider bootstraps the substrate of U1. With R_1 = 500 kΩ (source impedance), THD at 10 kHz was reduced an order of magnitude.

ELECTRONIC DESIGN

FIG. 57-40

SIMPLE SCHMITT TRIGGER

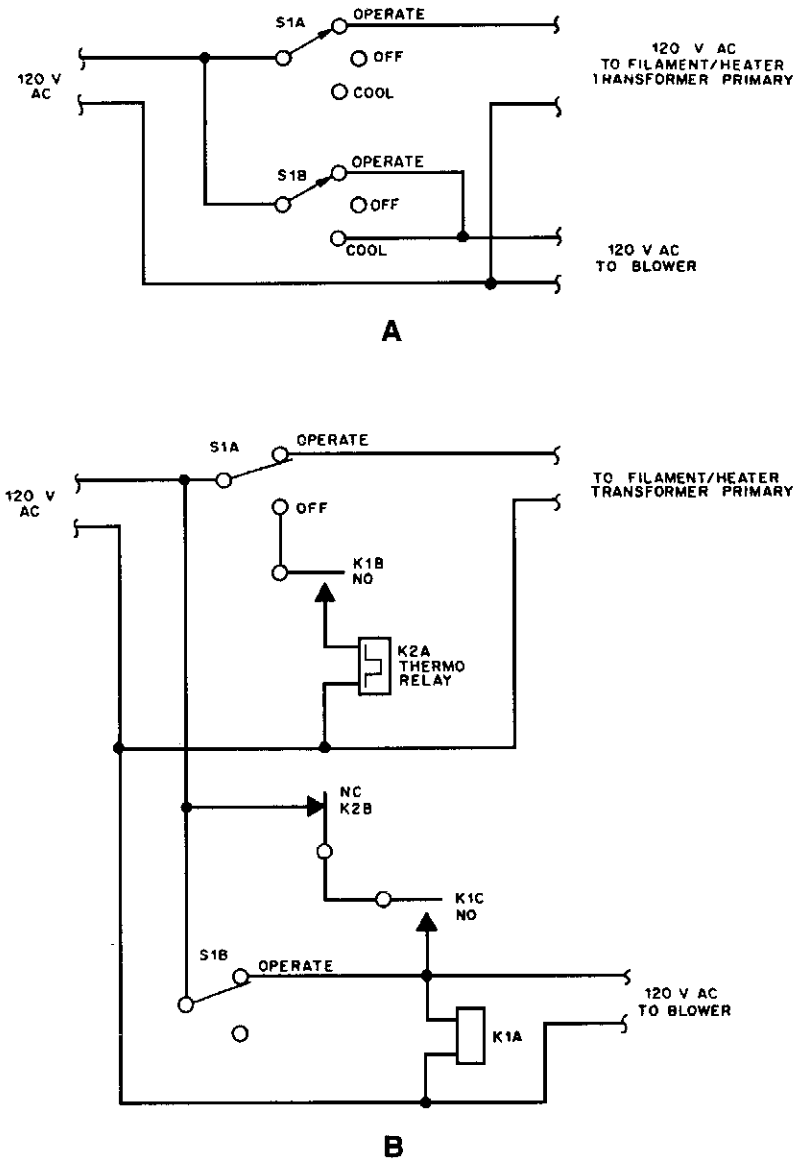


A 555 IC is shown configured to function as a Schmitt trigger. Inputs above and below the threshold level will turn the circuit on and off producing a square wave output.

WILLIAM SHEETS

FIG. 57-41

AMPLIFIER COOL-DOWN CIRCUIT II

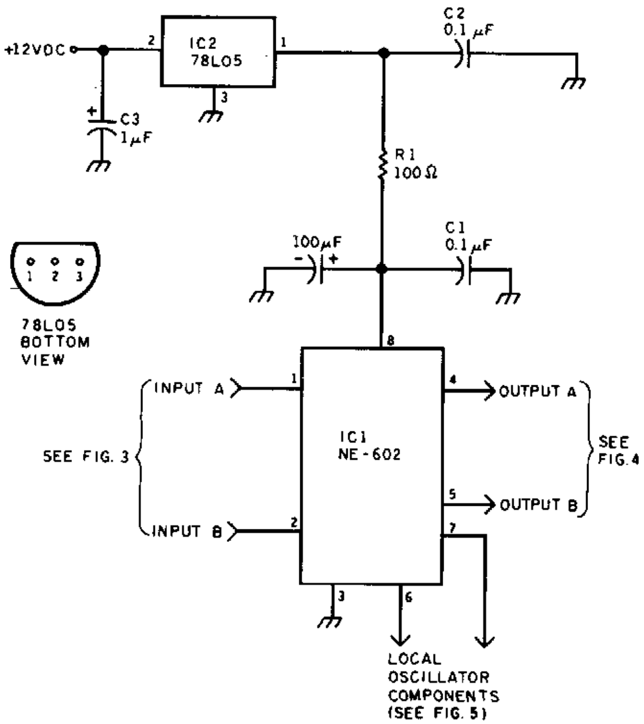


QST

FIG. 57-42

High-power amplifiers used in RF service, using vacuum tubes, often benefit from leaving the blower air flow on after removal of filament/heater voltage.

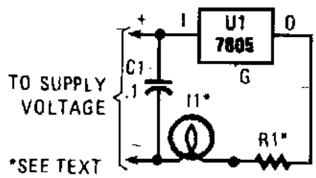
NE602 dc POWER CIRCUITS



73 AMATEUR RADIO TODAY ***FIG. 57-43***

The dc power supply circuit for the NE-602.

INRUSH CURRENT LIMITER



POPULAR ELECTRONICS ***FIG. 57-44***

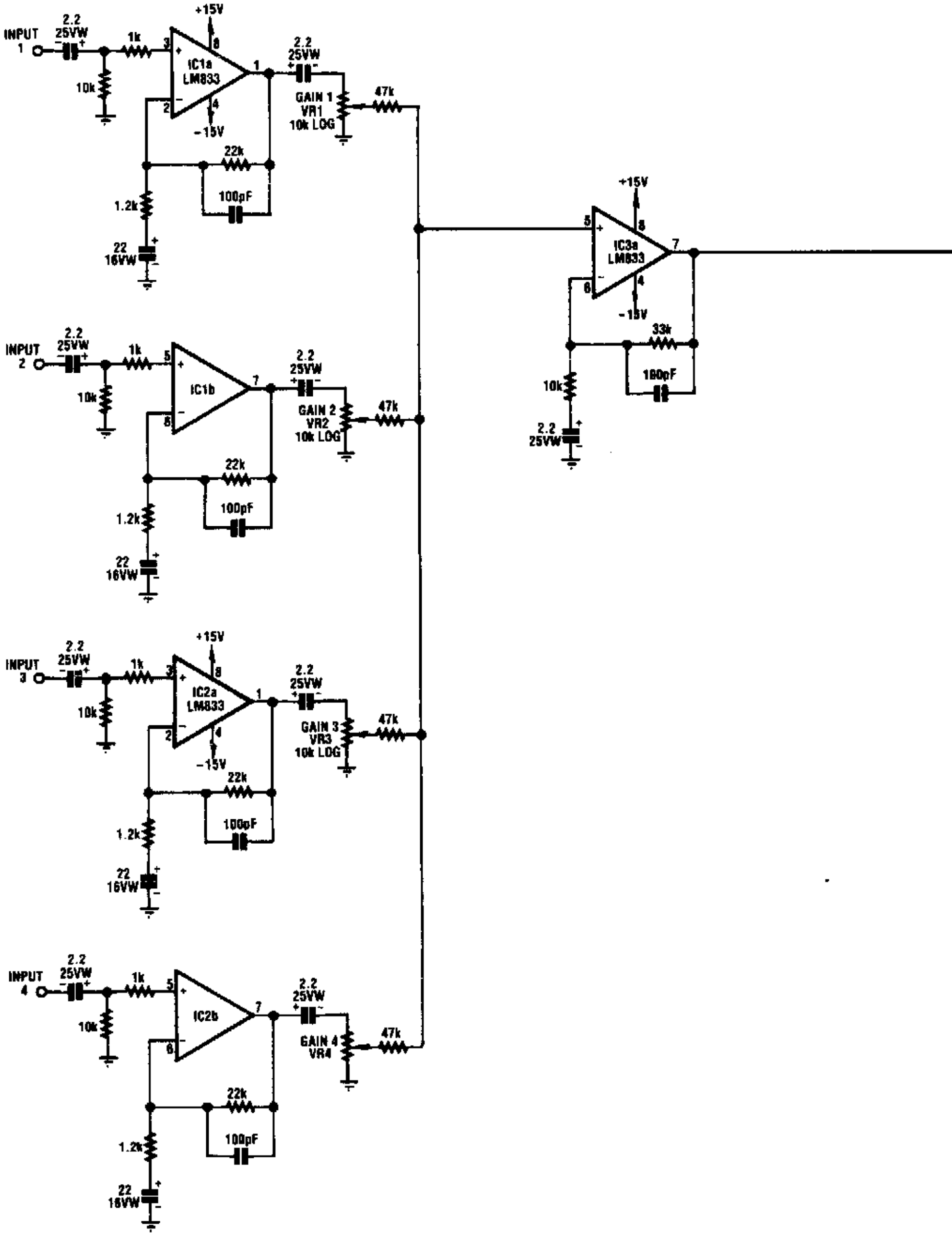
A 7805 can be configured as a constant-current regulator, to serve as an inrush current limiter. R1 will have 5 V across it at all times so the total current through I1 will be 5 V/R_1 + 5 mA, the 5 mA being the regulator operating current. In this case, R_1 = 5 V/95 mA = 52.6 Ω for I1 current = 100 mA.

58

Mixer Circuits

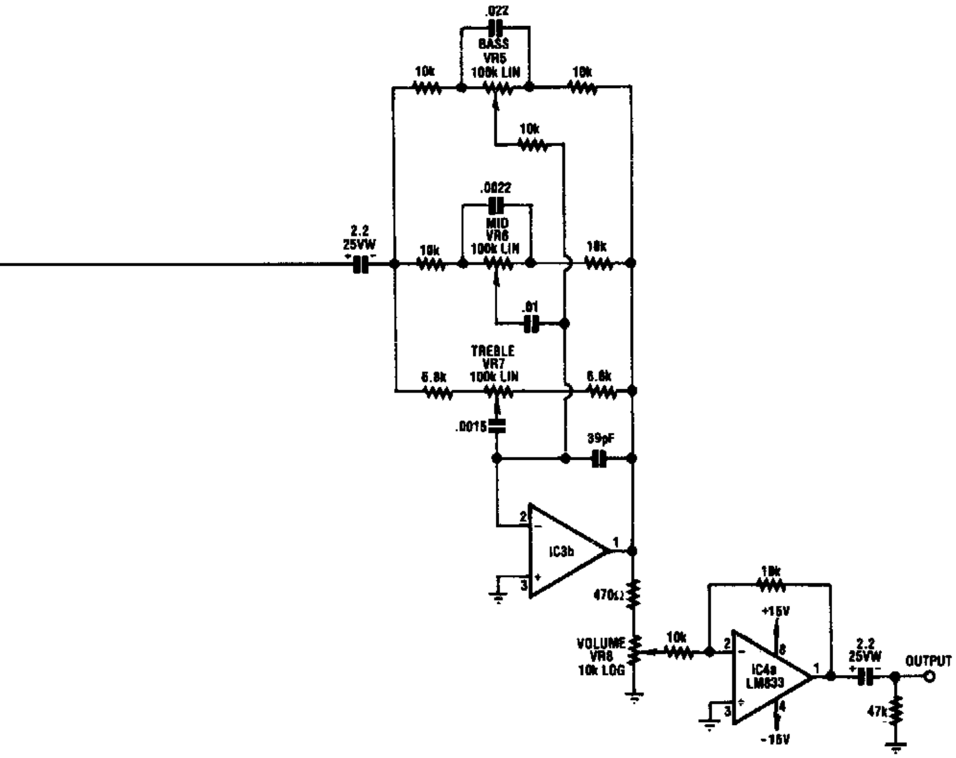
The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOW-NOISE 4-CHANNEL GUITAR MIXER

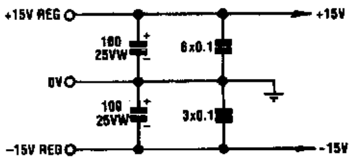


SILICON CHIP

IC1-a, IC1-b, IC2-a, and IC2-b all function with a gain of about 19. Their outputs are mixed via the level-control pots and the resulting signal amplified by IC3-a and fed to tone-control stage IC3-b. Finally, the output from IC3-b is fed to unity-gain buffer stage IC4-a via volume-control potentiometer VR8.



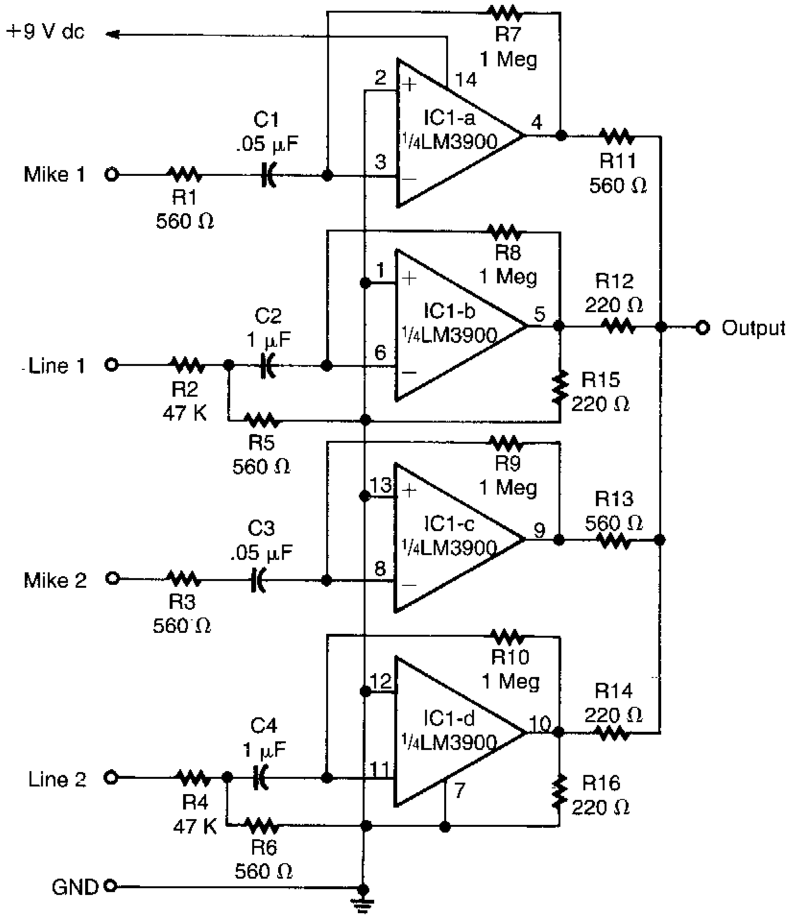
.022
BASS
VR5
100k LIN
10k
10k
10k
.0022
MID
VR6
100k LIN
2.2
25VW
10k
10k
.01
TREBLE
VR7
100k LIN
6.8k
6.8k
.0015
39pF
IC3b
470Ω
18k
+15V
VOLUME
VR8
10k LOG
10k
IC4a
LM833
2.2
25VW
OUTPUT
47k
-15V



+15V REG
+15V
100
25VW
6x0.1
0V
100
25VW
3x0.1
-15V REG
-15V

FIG. 58-1

AUDIO MIXER

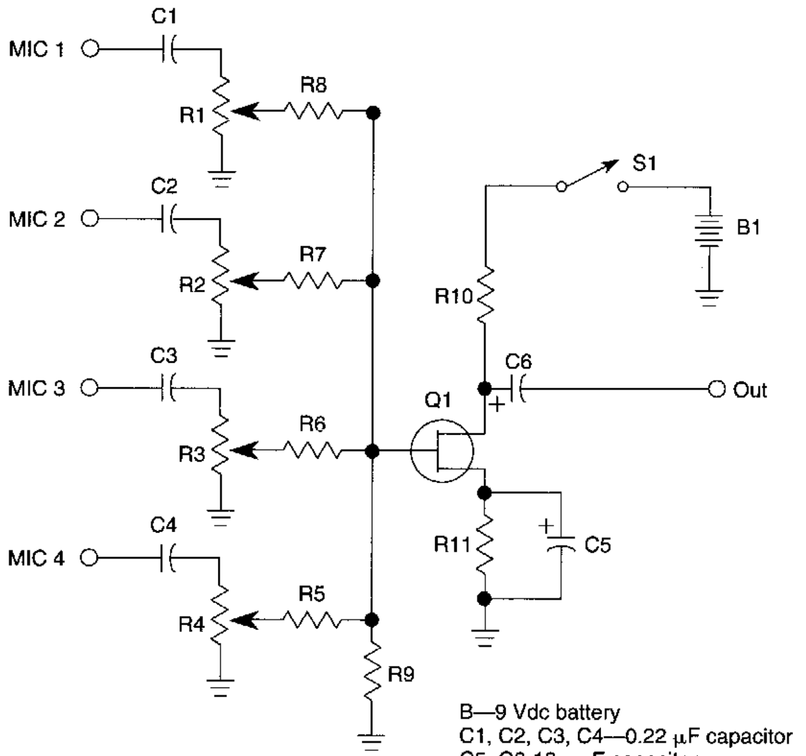


ELECTRONICS NOW

FIG. 58-2

Designed around an LM3900 quad op amp, this mixer combines 2-line and 2-mike inputs and sums them at the output terminal. R7 through R10 can be changed to vary the gain (around +23 dB).

FET MICROPHONE MIXER

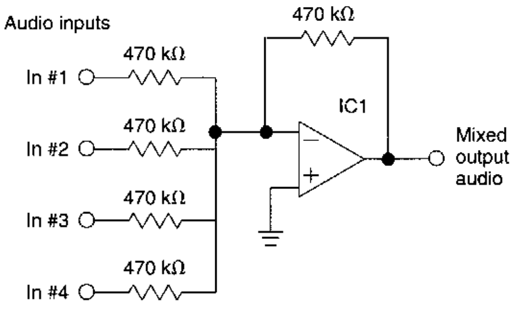


B—9 Vdc battery
C1, C2, C3, C4—0.22 μF capacitor
C5, C6-10—μF capacitor
R1, R2, R3, R4—500 kΩ potentiometer, audio taper
R5, R6, R7, R8—2.2 MΩ resistor
R9-10 MΩ resistor
R10—2.2 kΩ resistor
R11—470 Ω resistor
Q1—FET (field effect resistor), MPF102
S1—SPST switch

WILLIAM SHEETS ***FIG. 58-3***

A JFET transistor is used as a high-to-low impedance converter and signal mixer. Input impedance is approximately 500 kΩ but it can be increased by increasing R5 to R8 as high as 10 MΩ. Output Z is about 2 kΩ, but it can be increased or decreased by changing the value of R_{10}. Use 560 or 680 Ω to feed a 600-Ω input; use 100 kΩ to 1 MΩ for high impedance.

UNITY-GAIN FOUR-INPUT AUDIO MIXER



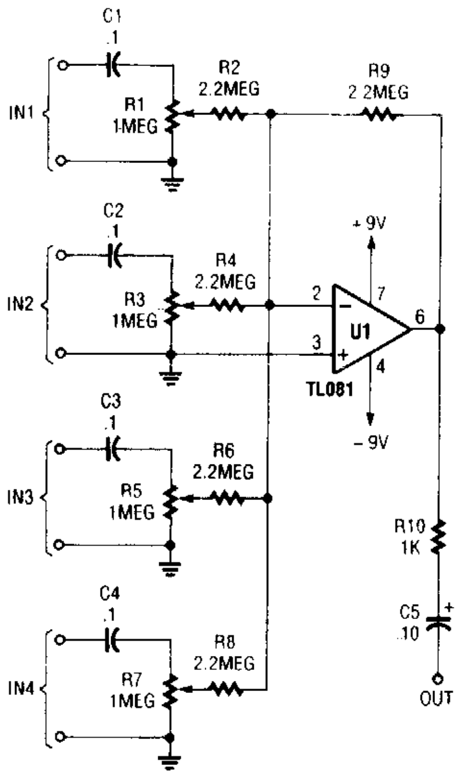
$E_{OUT} = -(In\ \#1 + In\ \#2 + In\ \#3 + In\ \#4)$

IC1 = LM741, etc.

The circuit has four inputs. The voltage gain between each input and the output is held at unity by the relative values of the 470kΩ input resistor and the 470kΩ feedback resistor.

WILLIAM SHEETS *FIG. 58-4*

FET OP AMP MICROPHONE MIXER



POPULAR ELECTRONICS *FIG. 58-5*

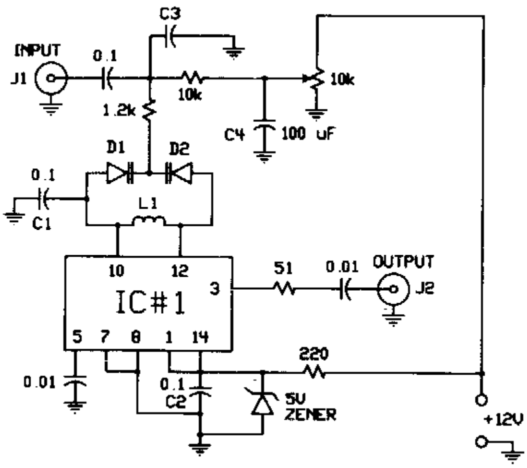
59

Modulator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

FM MODULATOR

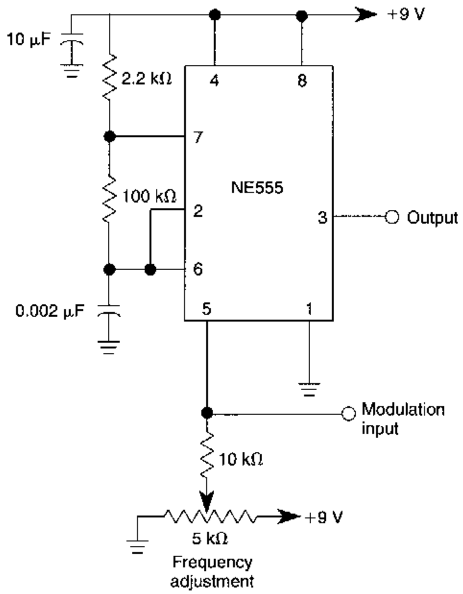
IC-1 - Motorola MC-1648P
All resistors 5%, 0.25 W
Zener - 5.1 V, 0.5 W
All 0.1 and 0.01 uF capacitors ceramic, 16V
C4 - 100 uF, 16 V electrolytic
D1, D2 - Motorola MV-209
L1 - airwound, 6 turns, 3/16" dia., 5/16" long, 20 AWG
C3 - 500 pF, silver mica



The FM modulator is built with a Motorola MC1648P oscillator. Two varactors, Motorola MV-209, are used to frequency modulate the oscillator. The 5000-Ω potentiometer is used to bias the varactors for best linearity. The output frequency of approximately 100 MHz can be adjusted by changing the value of the inductor. The output frequency can vary as much as 10 MHz on each side. The output level of the modulator is –5 dBm. In this prototype, the varactor bias was 7.5 V for best linearity; but this could be different with other varactors.

RF DESIGN *FIG. 59-1*

455-kHz MODULATOR

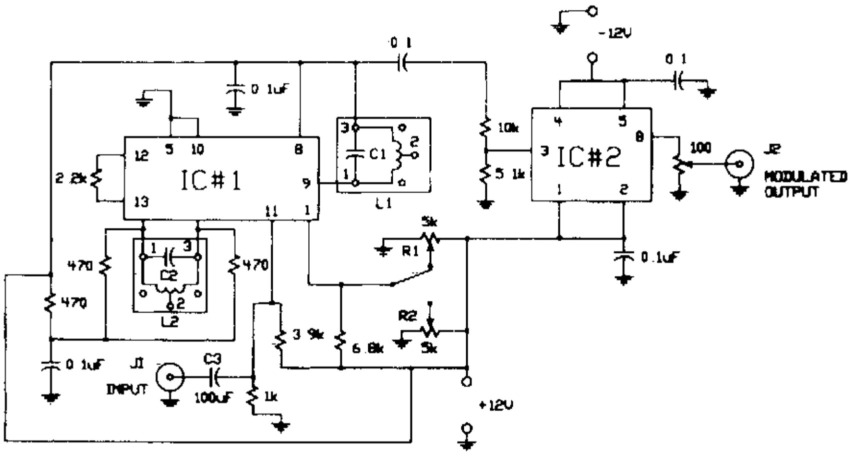


This circuit shows how to frequency-modulate the oscillator using a 555. Oscillator frequency is set with the 5-kΩ potentiometer and the modulation signal is dc-coupled.

WILLIAM SHEETS *FIG. 59-2*

555 FM CIRCUIT

IC-1 - Motorola MC-1374P
IC-2 - National LH0002C
L1, L2 - Mouser Electronics #421IF200
C1, C2 - silver mica, 300 pF
All 0.1 uF cap., ceramic disc, 16V
C3 - 100 uF, 10 V, electrolytic
All resistors 5%, 0.25 W
ADJUSTMENT: Adjust R1 for minimum carrier; signal from function generator should generate 500 mVpp at pin 8 of IC-2 (suppressed carrier double sideband). Adjust R2 and function generator level to achieve 800 mVpp at pin 8 of IC-2 (standard AM with carrier). Adjust L2 for 455 kHz. Adjust L1 for maximum output.



RF DESIGN

FIG. 59-3

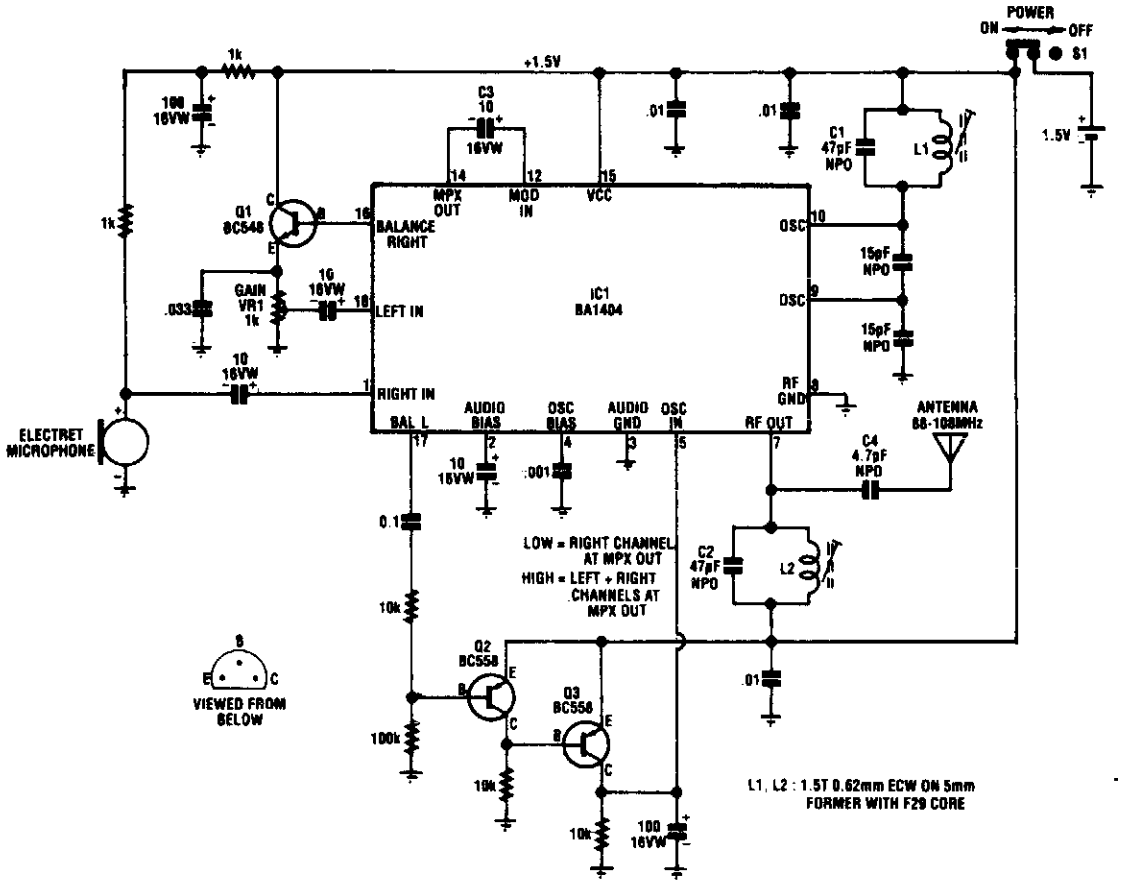
Circuit for applying a dc-coupled FM or PPM to a 555 configured as an oscillator.

60

Monitor Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ROOM MONITOR

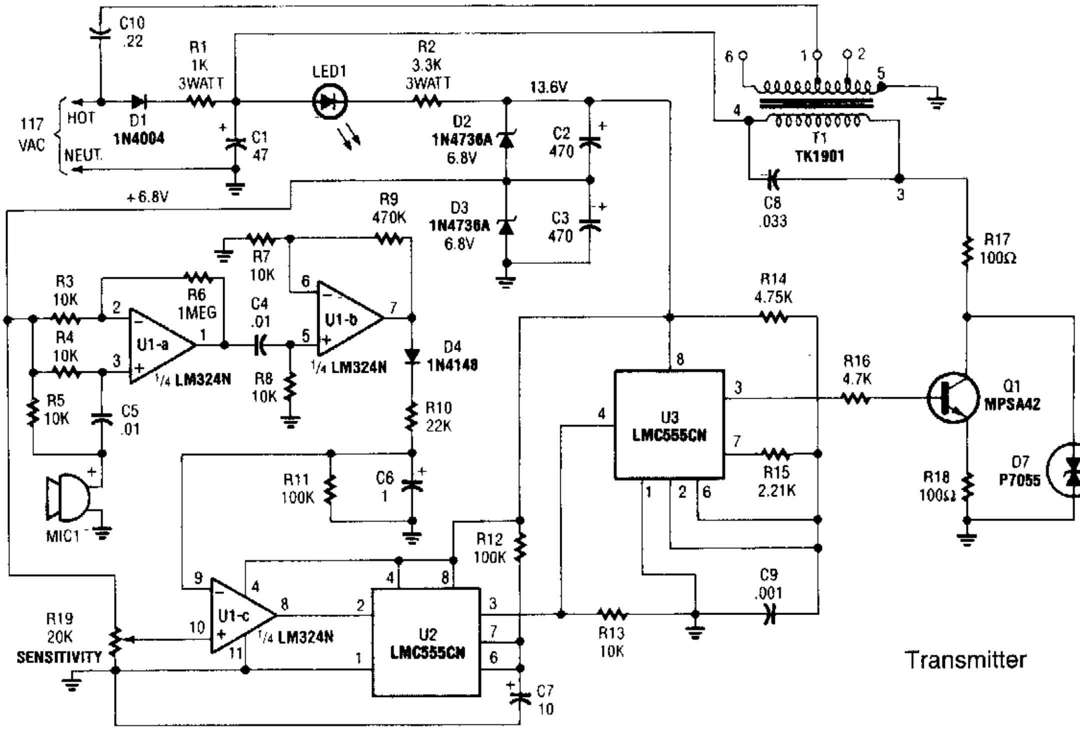


SILICON CHIP

FIG. 60-1

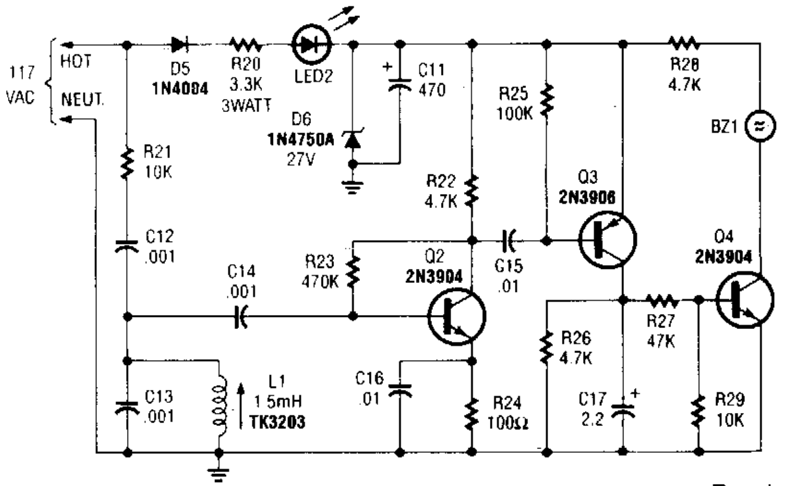
The circuit uses Q1 to buffer the right-channel balance output while Q2 and Q3 form a VOX circuit. When the signal level from the microphone goes high, the output of the VOX also goes high and the multiplexer inside IC1 switches the high-gain left-channel output through to a following buffer stage. This signal is then ac-coupled via C3 into an RF mixer stage and thence to an RF amplifier, which is tuned by C2 and L2.

BABY MONITOR



Transmitter

A



Receiver

B

FIG. 60-2

Transmitter operation. Operating power for the transmitter circuit is derived directly from the ac line. The dc power to operate the circuit is generated in two stages, one for an RF power-amplifier stage, and the second for the remainder of the circuit.

The ac line voltage is applied to D1, which half-wave rectifies the ac input. The resulting dc voltage (approximately 30V under load) is fed across an RC filter (comprised of R1 and C1) and used to operate amplifier, Q1. The second stage of the power supply (composed of LED1, R2, D2, D3, C2, and C3, which forms a regulated +13.6-V, center-tapped supply) feeds the remainder of the circuit. LED1 is connected in series with R2 and is used as a visual power-on indicator for the transmitter.

An electret microphone element (MIC1) is used as the pick-up. The output of the microphone is ac coupled through C5 to U1-a (a noninverting op amp with a gain of about 100). The output of U1-a at pin 1 is ac coupled through C4 to the noninverting input of U1-b (which provides an additional gain of 48) at pin 5. The output of U1-b at pin 7 is then fed through D4 and R10, and across R11 and C6 to the inverting input of U1-c which is biased to a positive voltage that is set by SENSITIVITY-control R19. This represents a threshold voltage at which the output of U1-c switches from high to low.

During standby, the output of U1-c at pin 8 is held at about 12 V when the voltage developed across C6 is less than the bias-voltage setting at pin 10. When a sound of sufficient intensity and duration is detected, the voltage at pin 9 of U1-c exceeds the threshold level (set by R19), causing U1-c's output at pin 8 at go low. That low is applied to pin 2 of U2 (a 555 oscillator/timer configured as a monostable multivibrator). This causes the output of U2 to go high for about one second, as determined by the time constant of R12 and C7. The output of U2 at pin 3 is applied to pin 4 of U3 (a second 555 oscillator/timer that is configured for astable operation, with a frequency of about 125 kHz). That causes U3 to oscillate, producing a near square-wave output that is used to drive Q1 into conduction. The output of Q1 is applied across a parallel-tuned circuit composed a T1's primary and C8. The tuned circuit, in turn, reshapes the 125-kHz signal, causing a sine-wave-like signal to appear across both the primary and the secondary of T1.

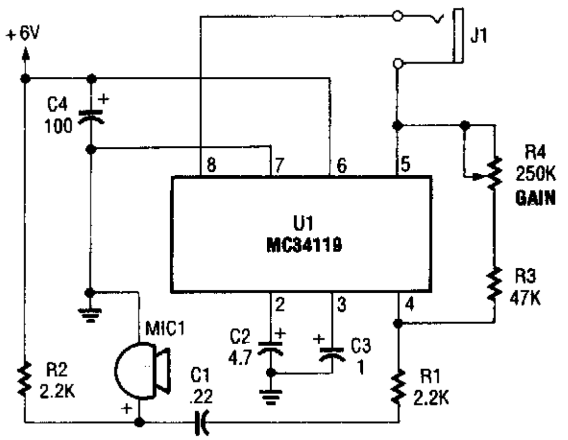
The signal appearing at T1's secondary (about 1 or 2 V peak-to-peak) is impressed across the ac power line, and is then distributed throughout the building without affecting other electrical appliances connected to the line. Transient suppressor D7 is included in the circuit to help protect Q1 from voltage spikes that might appear across the power line and be coupled to the circuit through T1.

Receiver operation. Power for the receiver, as with the transmitter, is derived from a traditional half-wave rectifier (D5). The resulting dc voltage is regulated to 27 V by D6 and R20, and is then filtered by C11 to provide a relatively clean, dc power source for the circuit. A light-emitting diode, LED2, connected in series with R20 provides a visual indication that the circuit is powered and ready to receive a signal.

The 125-kHz signal is plucked from the ac line and coupled through R21 and C12 to a parallel-tuned LC circuit, consisting of C13 and L1. That LC circuit passes 125-kHz signals while attenuating all others. The 125-kHz signal is fed through C14 to the base of Q2 (which is configured as a high-gain linear amplifier), which boosts the relatively low amplitude of the 125-kHz signal. The RF output of Q2 is ac coupled to the base of Q3 through C15. Transistor Q3 acts as both an amplifier and detector. Because there is no bias voltage applied to the base of Q3, it remains cut off until driven by the amplified 125-kHz signal. When Q3 is forward biased, its collector voltage rises.

Capacitor C16, connected across Q3's collector resistor, filters the 125-kHz signal so that it is essentially dc. When the voltage at the collector of Q3 rises, Q4 is driven into conduction. That causes current to flow into piezo buzzer BZ1, producing a distinctive audio tone that alerts anyone within earshot that the baby needs attention.

BIRD FEEDER MONITOR



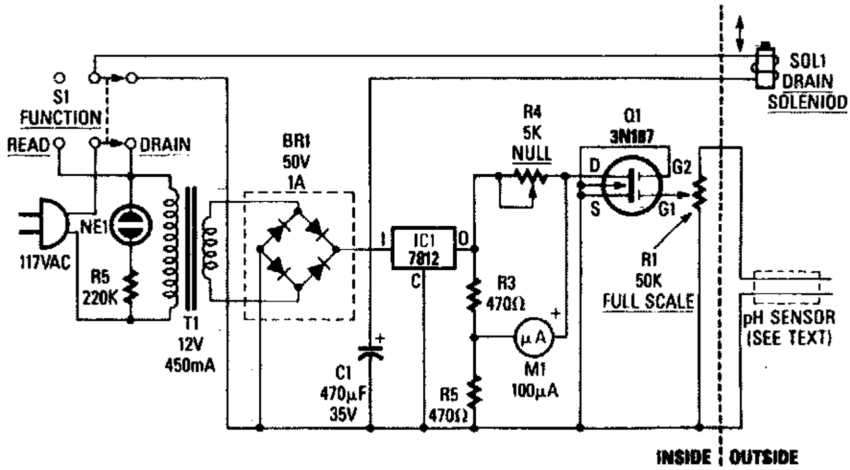
POPULAR ELECTRONICS

FIG. 60-3

The first amplifier circuit is a bird phone. In this circuit, the electret mike (MIC1) is mounted in the neck of a large plastic funnel. The amplifier, built around an MC34119 (which is available from D.C. Electronics, P.O. Box 3203, Scottsdale, AZ 85271-3203; Tel. 800-467-7736, and elsewhere), is then placed outside of the funnel with the pick-up facing a nearby bird feeder. The output of the amplifier is then connected to a 16-Ω speaker.

The amplifier's voltage gain is determined by the values of the input resistor (R1) and the feed-back resistor (R3 and R4, respectively). The differential gain of the amplifier is given by: $R_3 + R_4/R_1 \times 2$. With the component values shown, the maximum voltage gain is about 270. This permits listening to the activity at the bird feeder.

ACID-RAIN MONITOR



R-E EXPERIMENTERS HANDBOOK

FIG. 60-4

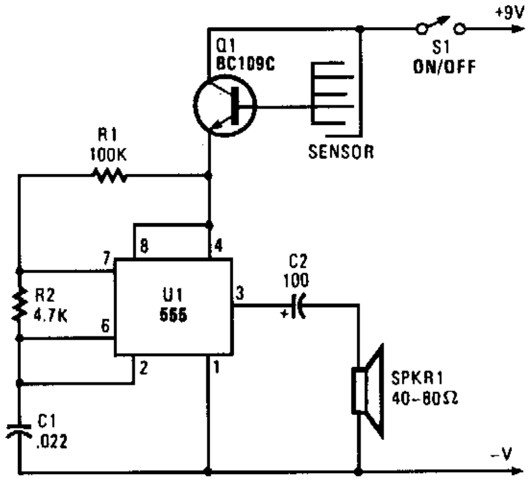
The drain-to-source resistance of Q1 varies depending on the acidity of the sample presented to Q1's gate circuit. That variable resistance varies the current flowing through the bridge; that current is proportional to pH.

61

Moisture- and Fluid-Detector Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

WATER-ACTIVATED ALARM

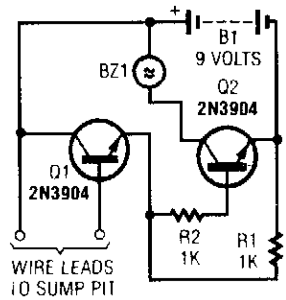


POPULAR ELECTRONICS

FIG. 61-1

When sensor gets wet, it conducts, forward-biases Q1, and activates audio oscillator U1. A tone is heard from the speaker.

SIMPLE FLOOD ALARM

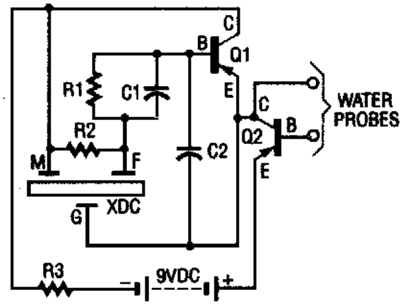


POPULAR ELECTRONICS

FIG. 61-2

A common collector amplifier drives a 2N3904 switch to sound alarm BZ1. The wire leads to water sensor or sump pit, level switch, etc. and used to allow the alarm to operate and be mounted in a dry place.

MOISTURE DETECTOR



1991 PE HOBBYIST HANDBOOK ***FIG. 61-3***

The moisture detector uses two transistors and a piezoelectric transducer to sound an alarm tone when water is present. Transistor Q1 forms a crystal-controlled oscillator, using a portion of piezoelectric transducer XDC—which contains two piezoelectric crystal regions—as the crystal. The transducer has three separate leads. One lead goes to each of the crystals, and the third lead is common to both.

The smaller internal crystal region sets the frequency of operation and the larger element is driven by Q1 (when it is biased "on") to provide the loud tone output. To turn the pnp transistor Q1 (used as an oscillator) "on" pnp transistor Q2 (used here as a switch) must be on. To turn it "on" with the biasing that is normally connected, you would only need to connect a resistor from the collector of Q2 to the base, which gives the base a negative (–) bias. The resistor used is the water that is to be detected. That turns Q2 on, which, in turn, turns on Q1. The result when water touches the probe is that the transducer emits a loud sound.

C1, C2	0.1-µF Mylar Capacitor
Q1, Q2	2N3906 Transistor
R1	6.8-kΩ Resistor
R2	33-kΩ Resistor
R3	200-Ω Resistor
XDC	Piezoelectric Transducer

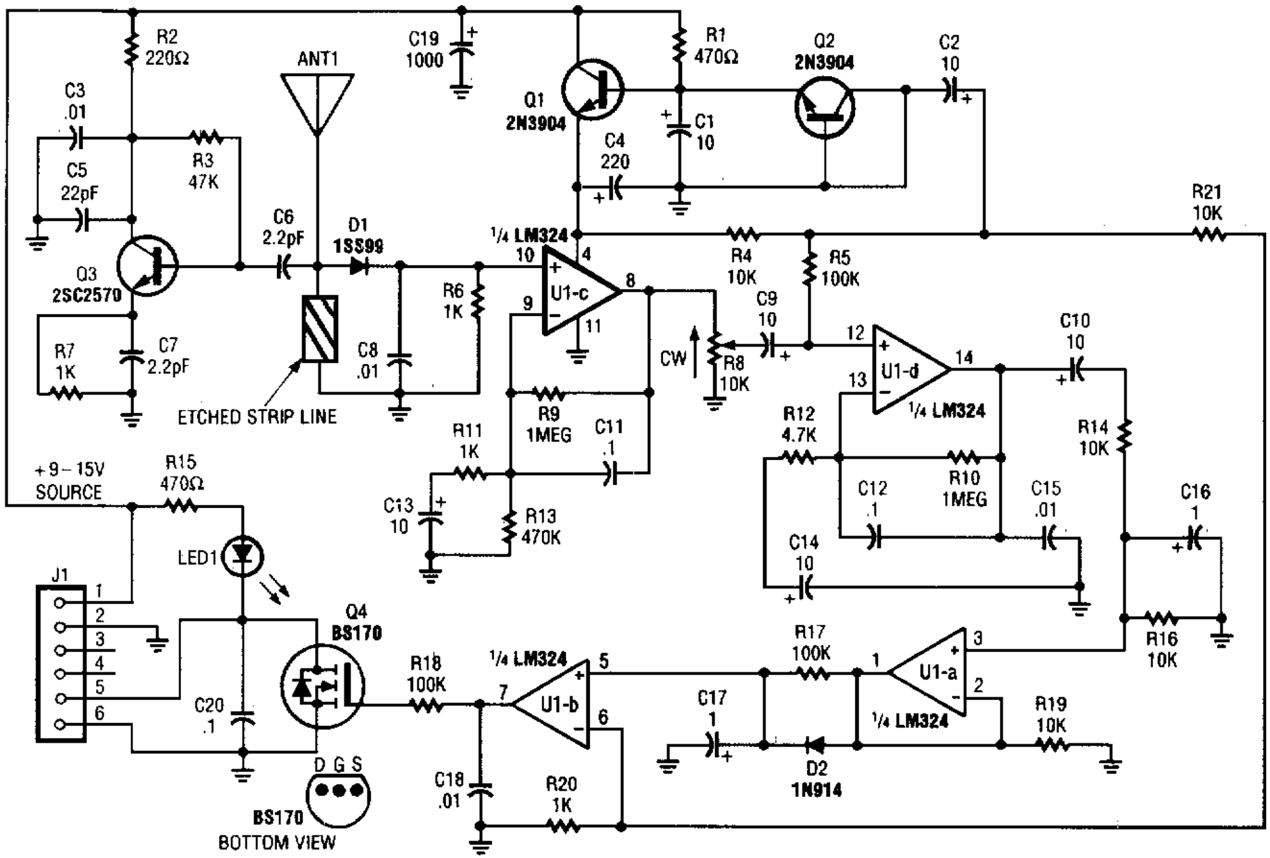
62

Motion Detector Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Microwave Motion Detector

MICROWAVE MOTION DETECTOR



POPULAR ELECTRONICS

FIG. 62-1

Operating at around 1.1 GHz, the detector senses field disturbance in the neighborhood of the antenna. The Doppler signal from detector D1 is amplified and drives a power MOSFET switch. The antenna is a short (2 to 3") length of wire.

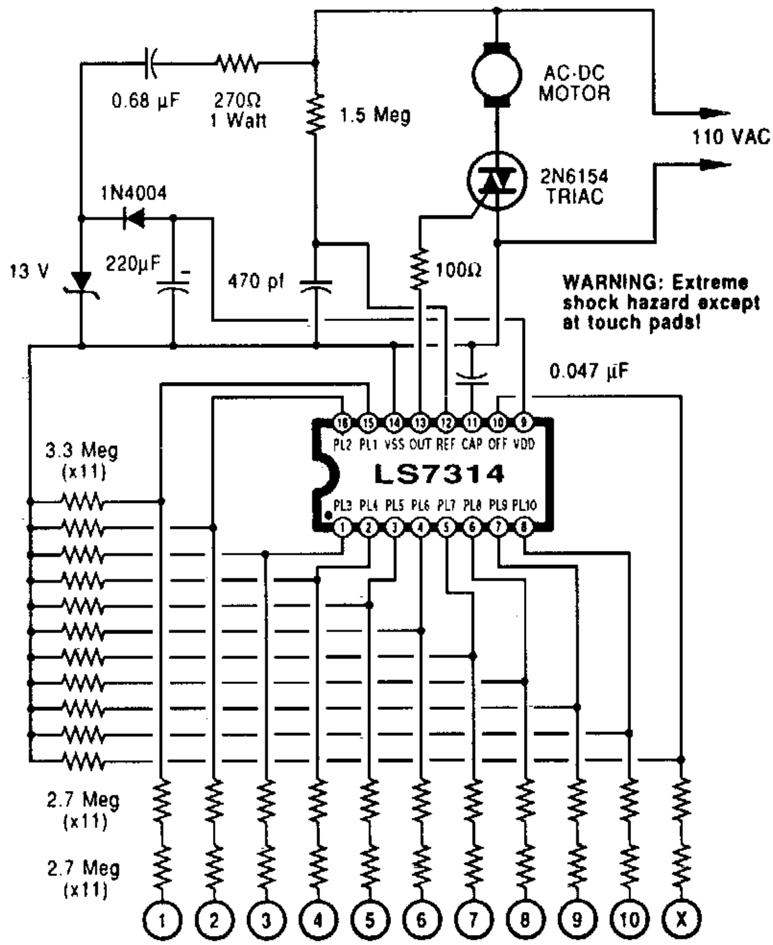
63

Motor-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Blender-Control Circuit
PWM Motor-Drive Circuit
Speed-Control Switch Circuit
Piezo Motor Drive
Pulse-Width-Modulated Motor-Speed Control
Speed-Control Switch

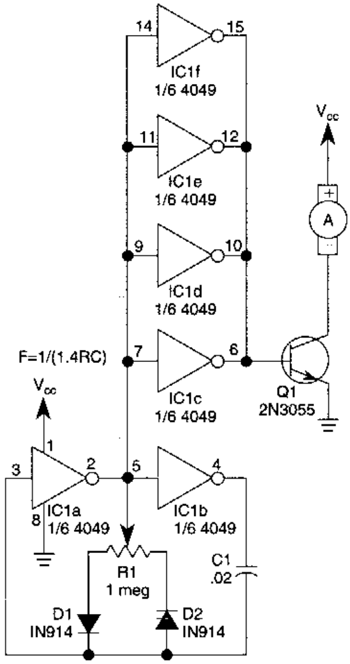
BLENDER-CONTROL CIRCUIT



RADIO-ELECTRONICS *FIG. 63-1*

A 10-speed touch-control blender circuit that uses the low-cost LS314 chip by LSI Systems. The 11th touch pad is for power off.

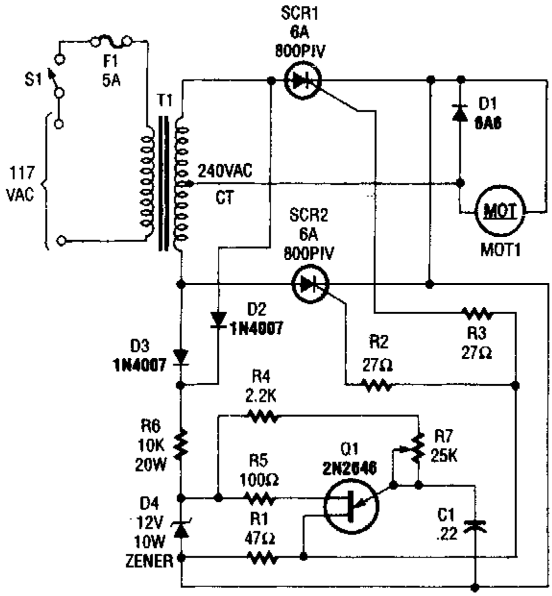
PWM MOTOR-DRIVE CIRCUIT



RADIO-ELECTRONICS **FIG. 63-2**

This circuit will drive a small dc motor over a wide range of speeds without stalling by controlling the duty cycle of the motor, rather than the supply voltage.

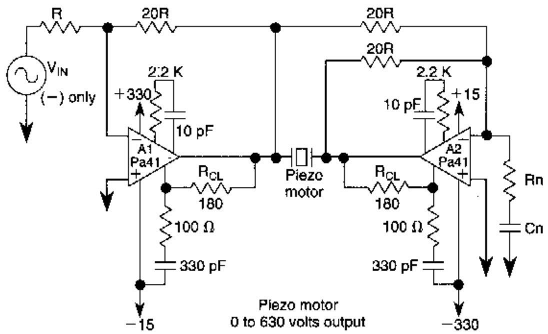
SPEED-CONTROL SWITCH CIRCUIT



POPULAR ELECTRONICS **FIG. 63-3**

A center-tapped 240-V transformer is used with two SCR devices to provide rectified ac (pulsating dc) to MOT1. Q1 is a UJT ramp generator used to generate trigger pulses for SCR1 and SCR2.

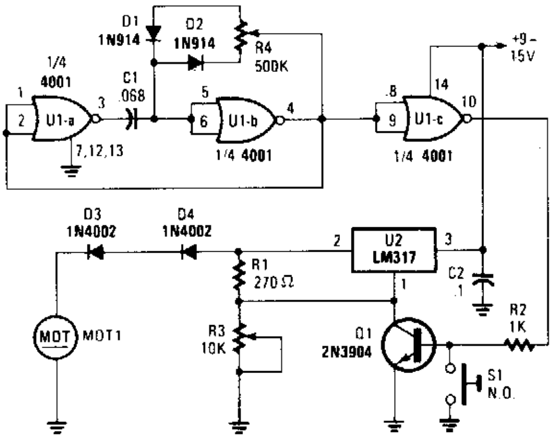
PIEZO MOTOR DRIVE



ELECTRONIC DESIGN **FIG. 63-4**

Using two Apex Microtechnology PA41 devices in a bridge circuit, this piezo motor driver delivers 0- to 630-V output.

PULSE-WIDTH-MODULATED MOTOR-SPEED CONTROL



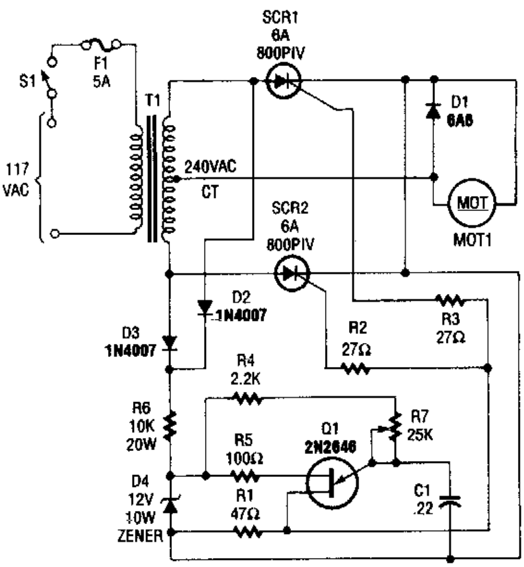
POPULAR ELECTRONICS

FIG. 63-5

Connected in this manner, an LM317 1-A adjustable-voltage regulator can be used to control the speed of a miniature dc motor or vary the brilliance of a small lamp. The circuit does so by controlling the pulse width, and therefore the current, to the load device.

To set the desired maximum output voltage, momentarily close S1 and adjust R3. Connect either a lamp or small dc motor (as is shown in the schematic to the circuit's output) and adjust R4 for the desired results. Any device that is driven by this circuit should have a current requirement of 1 A or less. And you should be sure to use good-sized heatsink for the LM317 regulator IC.

SPEED-CONTROL SWITCH



POPULAR ELECTRONICS

FIG. 63-6

The speed-control switch offers reasonably good control and stability to both ends of its operating range. This circuit uses two SCR devices in a full-wave configuration to control the dc power to a motor. A center-tapped transformer is used to supply the SCRs.

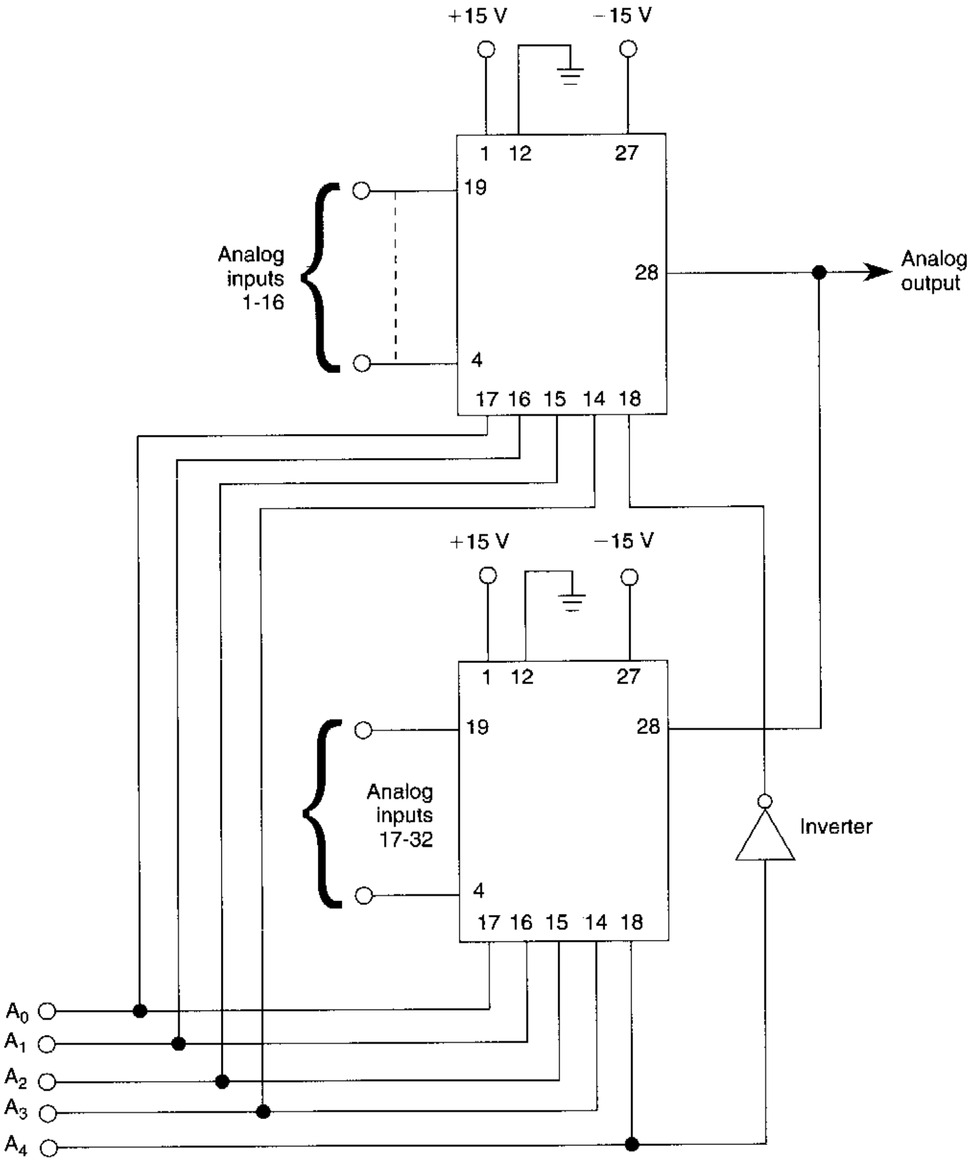
64

Multiplexer Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

32-Channel Analog Multiplexer

32-CHANNEL ANALOG MULTIPLEXER



WILLIAM SHEETS

FIG. 64-1

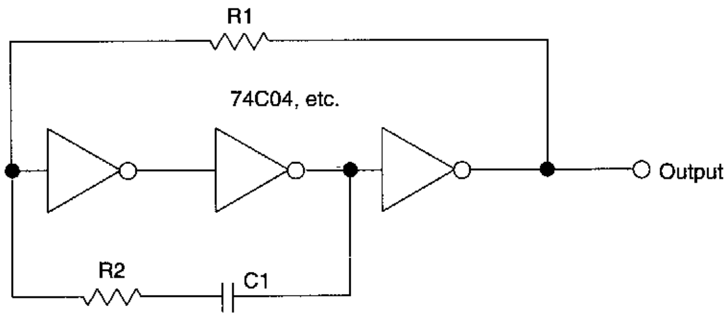
Using two Siliconix DG506 multiplexer chips, this 32-channel analog multiplexer selects 1 of 32 channels, depending on the data inputs $A_0 - A_4$.

65

Multivibrator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

IMPROVED CMOS MULTIVIBRATOR

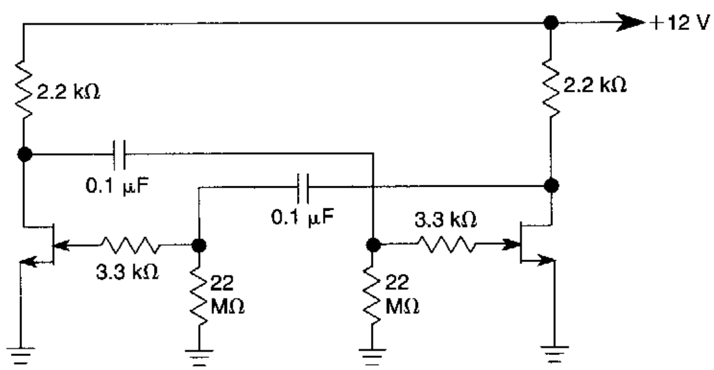


WILLIAM SHEETS

FIG. 65-1

This circuit uses a protective resistor $R2$ in conjunction with feedback resistor $R1$. Together, they form a voltage divider to reduce the input voltage amplitude for IC1-a so that the protective diodes never conduct. This improves temperature and voltage stability of the multivibrator.

VERY LOW FREQUENCY MULTIVIBRATOR



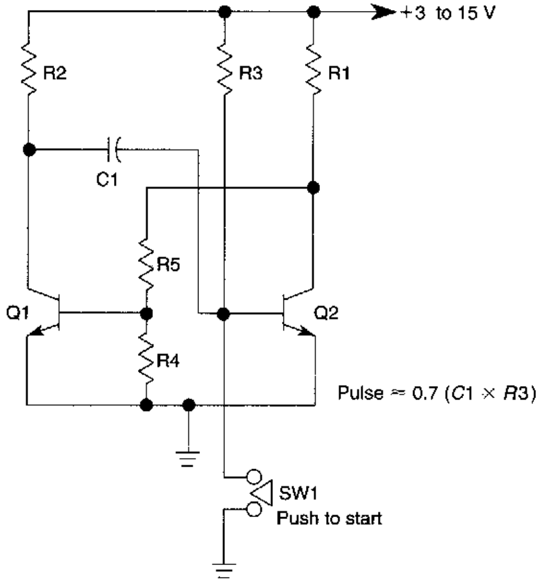
JFETs Transistor: N-channel (MPF102, etc.)

WILLIAM SHEETS

FIG. 65-2

The use of JFETs permits, high resistance and long time constants in this very low frequency multivibrator. The values shown are for 0.15 Hz operation.

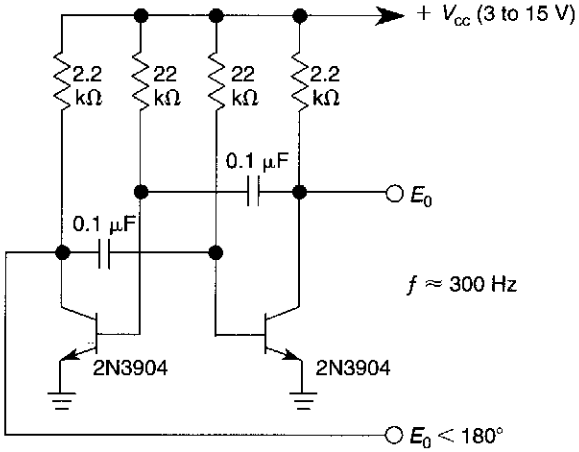
MONOSTABLE MULTIVIBRATOR I



WILLIAM SHEETS **FIG. 65-3**

This circuit is activated when SW1 is pushed to ground the base of transistor Q2. The pulse rate is approximately equal to 0.7(R3×C1).

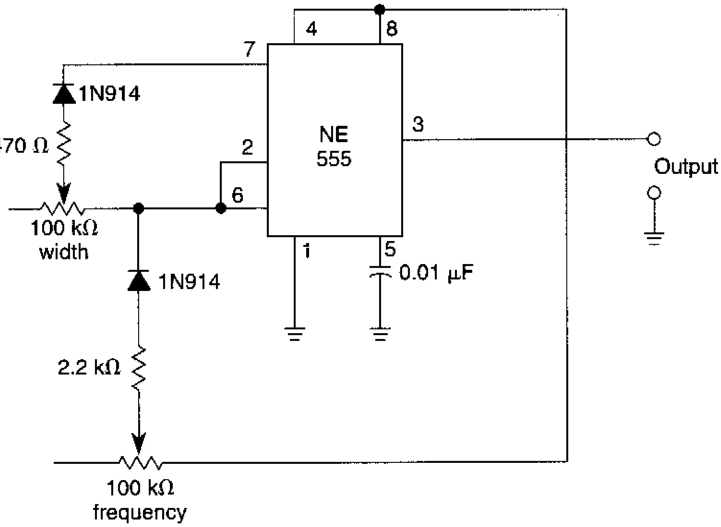
ASTABLE MULTIVIBRATOR OR FREE-RUNNING SQUARE-WAVE OSCILLATOR



WILLIAM SHEETS **FIG. 65-4**

This free-running square-wave oscillator uses two npn transistors. Output frequency is approximately 300 Hz with the values shown.

ASTABLE MULTIVIBRATOR I

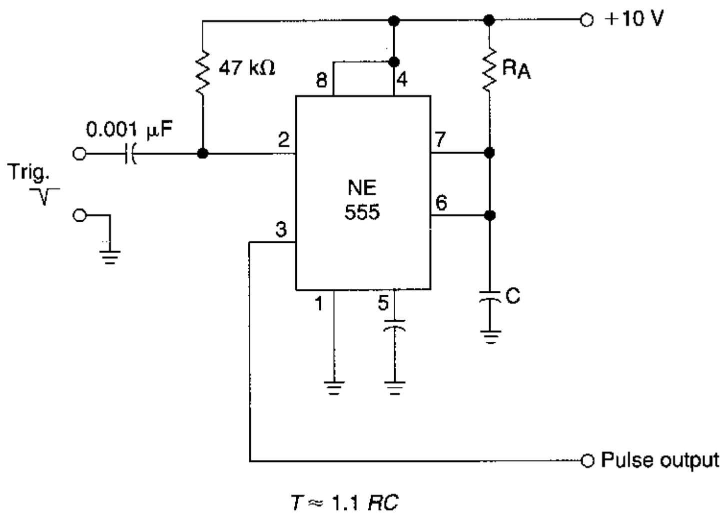


WILLIAM SHEETS

FIG. 65-5

In this multivibrator circuit frequency and pulse width can be separately controlled by using steering diodes (1N914) and two potentiometers.

MONOSTABLE MULTIVIBRATOR II

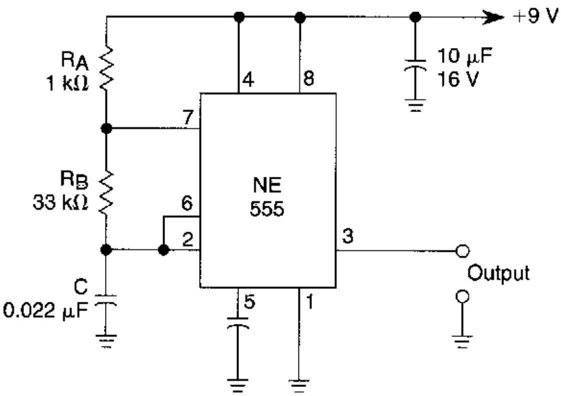


WILLIAM SHEETS

FIG. 65-6

The time constant of R_AXC determines the period of the monostable multivibrator. A negative pulse at pin 2 of the 555 starts the cycle.

ASTABLE MULTIVIBRATOR II

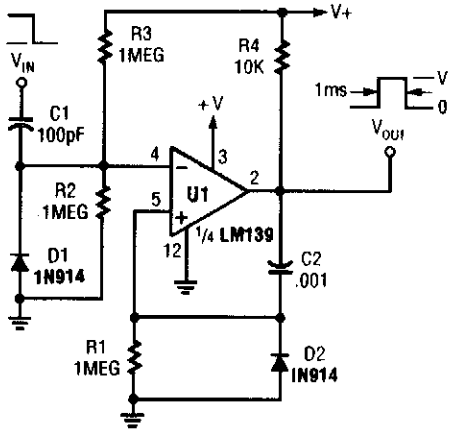


$$f = \frac{1.44}{(R_A + 2R_B)\,C}$$

WILLIAM SHEETS **FIG. 65-7**

An astable multivibrator based on the 555 is shown. Freq is approximately 975 Hz as determined by the values of R_B and C.

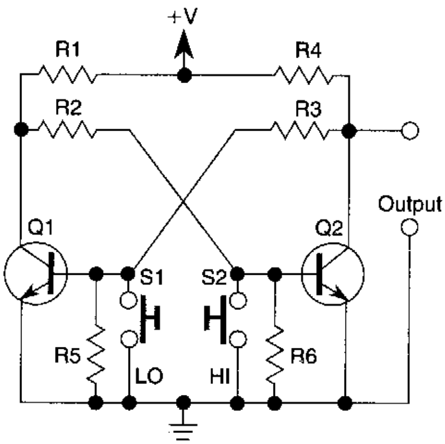
ONE-SHOT MULTIVIBRATOR



POPULAR ELECTRONICS **FIG. 65-8**

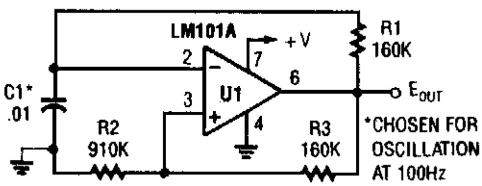
A section of a quad LM139 is used here as a one-shot pulse former.

FLIP-FLOP OR BISTABLE MULTIVIBRATOR WITH PUSHBUTTON TRIGGERING



ELECTRONICS NOW **FIG. 65-9**

FREE-RUNNING MULTIVIBRATOR USING OP AMP



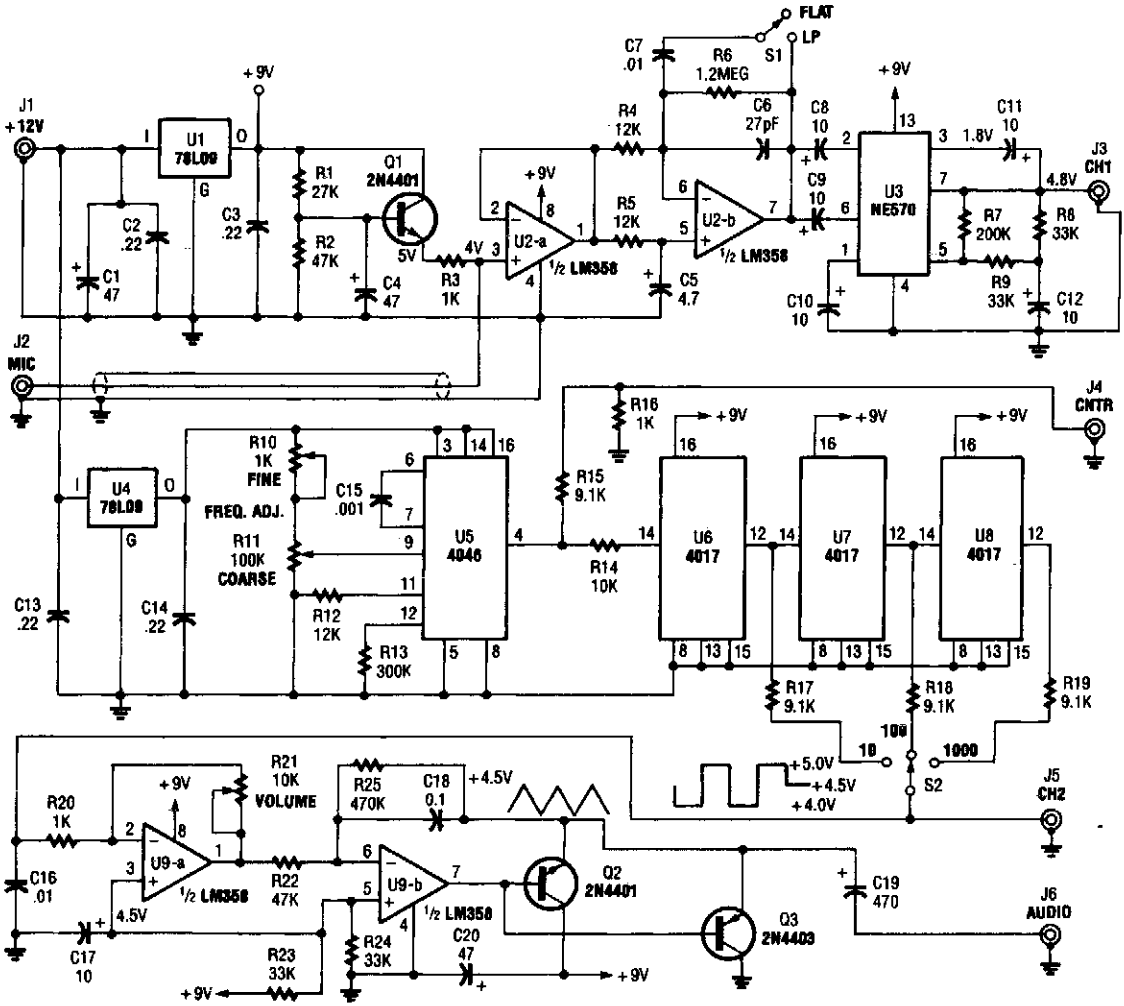
POPULAR ELECTRONICS **FIG. 65-10**

66

Musical Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 575. The figure number in the box of each circuit correlates to the entry in the Sources section.

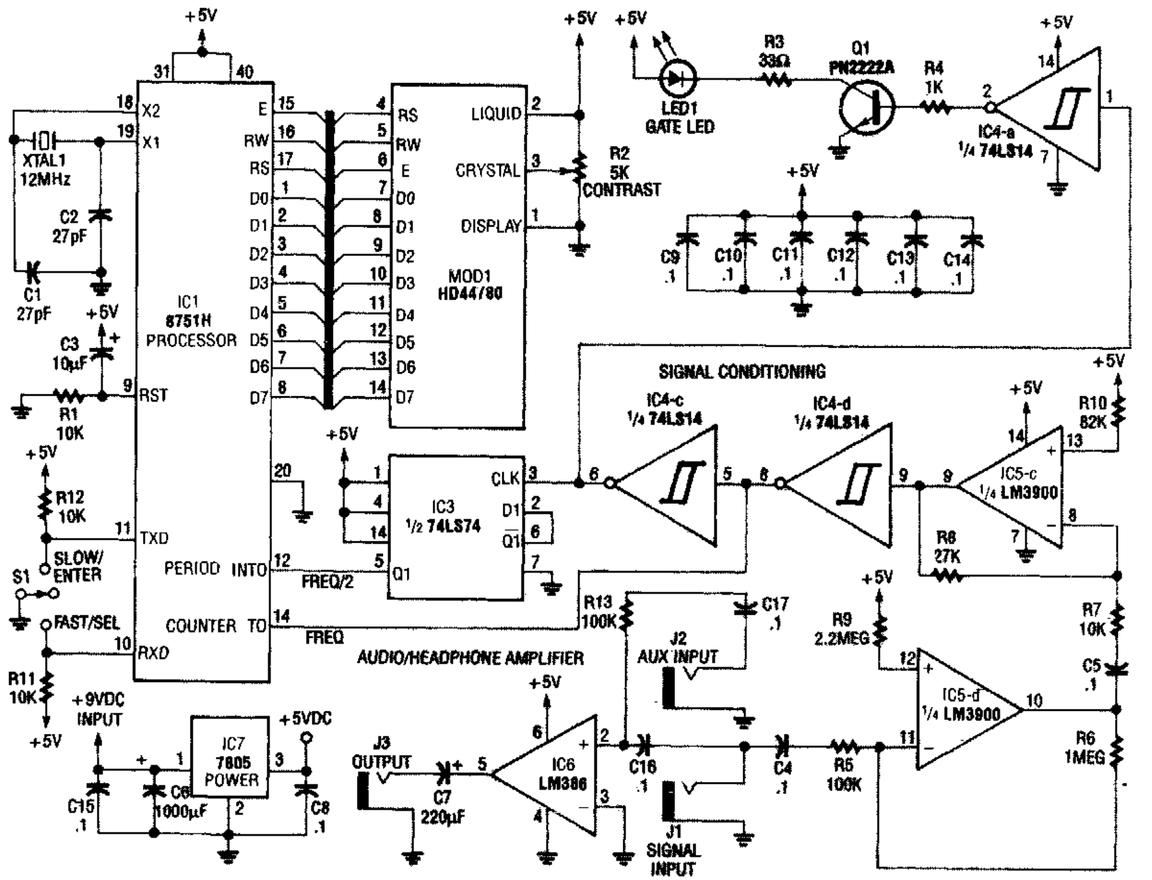
PRECISION AUDIO GENERATOR FOR MUSICAL INSTRUMENT TUNE-UP



1993 ELECTRONICS HOBBYIST HANDBOOK *FIG. 66-1*

One section of the precision audio frequency generator uses an electret microphone element to pick up audio from the piano. That signal is then processed and sent to one channel of a dual-trace oscilloscope. The other section of the circuit is used to produce a variable-frequency signal that is fed to a digital frequency counter. After conditioning, the audio signal is presented to the second channel of the scope and output to a set of stereo headphones.

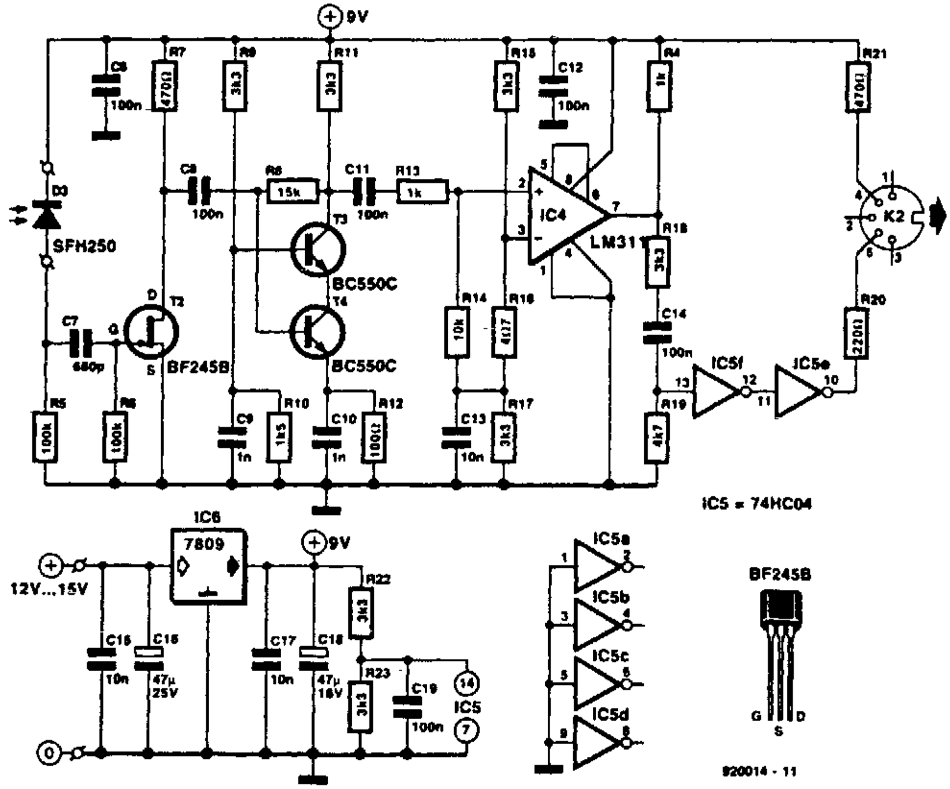
PERFECT PITCH



ELECTRONICS NOW *FIG. 66-2*

Perfect pitch, which is based on the 8751 H microprocessor, is an inexpensive and easy-to-build instrument tuner/frequency counter with a built-in headphone amplifier and a visual metronome. Perfect pitch converts the audio signal from your instrument to a digital signal, and displays the musical note you are playing and its frequency in real time on a 16-character liquid-crystal display. It also has an auxiliary audio input for radio, tape, or CD players so that you can tune up and play along with your favorite artists.

MUSICAL INSTRUMENT DIGITAL INTERFACE (MIDI) RECEIVER

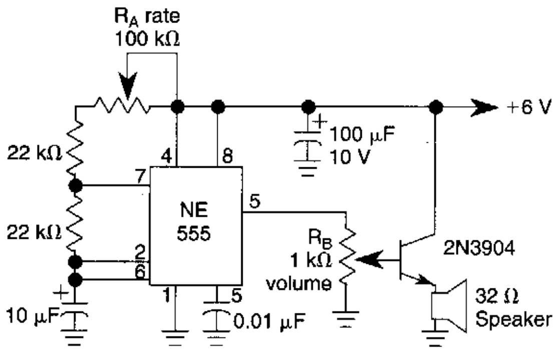


ELEKTOR ELECTRONICS

FIG. 66-3

Receiver photodiode SFH250 is used to convert optical data pulses at 32.5 Kb to electrical signals. Buffer T2 feeds the signals to cascade amplifier T3-T4, then to op amp IC4, and buffers IC5-f and IC5-e. IC6 supplies 9 V for the circuit.

ELECTRONIC METRONOME

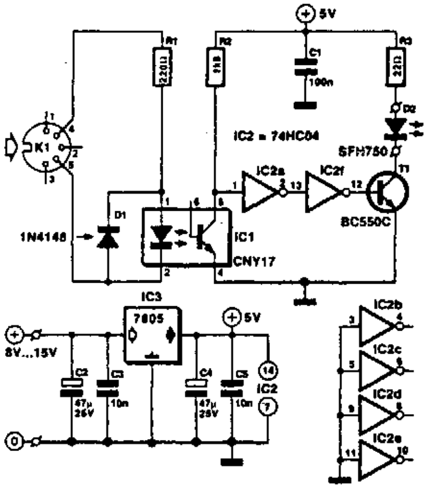


R_A sets the rate while R_B sets the volume of clocks in the speaker. The 555 is configured as a low frequency oscillator. The circuit is powered by a 6 V battery.

ELECTRONICS NOW

FIG. 66-4

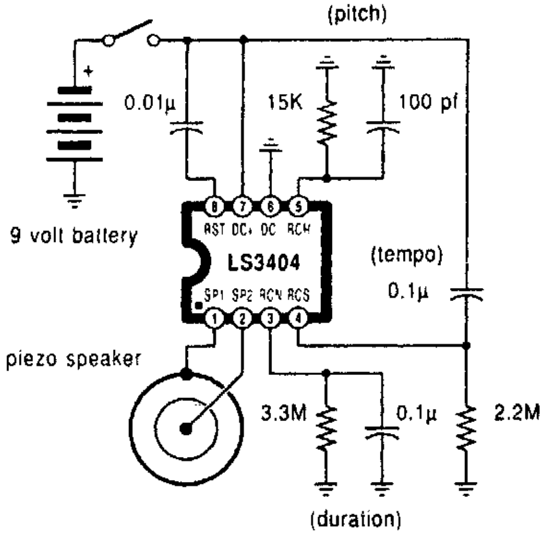
MUSICAL INSTRUMENT DIGITAL INTERFACE (MIDI) TRANSMITTER



ELEKTOR ELECTRONICS ***FIG. 66-5***

Used for digital control of musical instruments, this transmitter converts the digital data signals to equivalent optical signals for fiberoptic cable interface. Optocoupler IC1 provides isolation, and drives IC2-a and -b and T1, and finally provides a cable driver LED (SFH750).

MELODY CIRCUIT

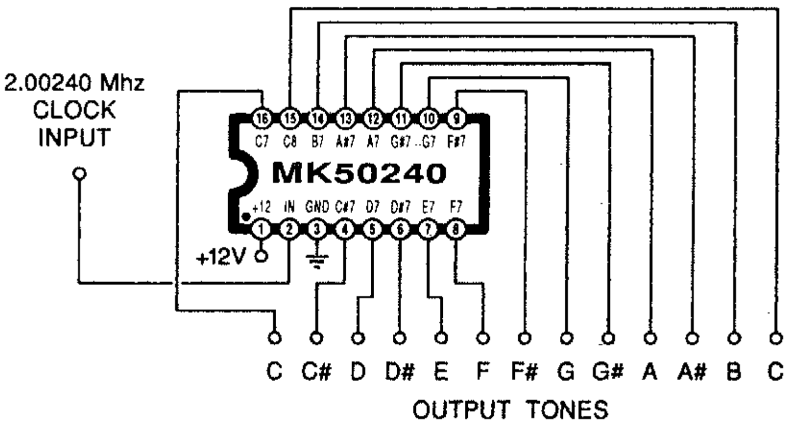


RADIO-ELECTRONICS ***FIG. 66-6***

A high-quality melody circuit. The slow decay waveform produced will create chime-like notes. Pitch, tempo, and duration are all adjustable.

TOP OCTAVE GENERATOR

Inputs and outputs are 12 volt square waves



RADIO-ELECTRONICS ***FIG. 66-7***

Using an MK50240, this circuit produces 12 top octave tones. The input and output lines can be divided using a binary divider IC to obtain the lower notes.

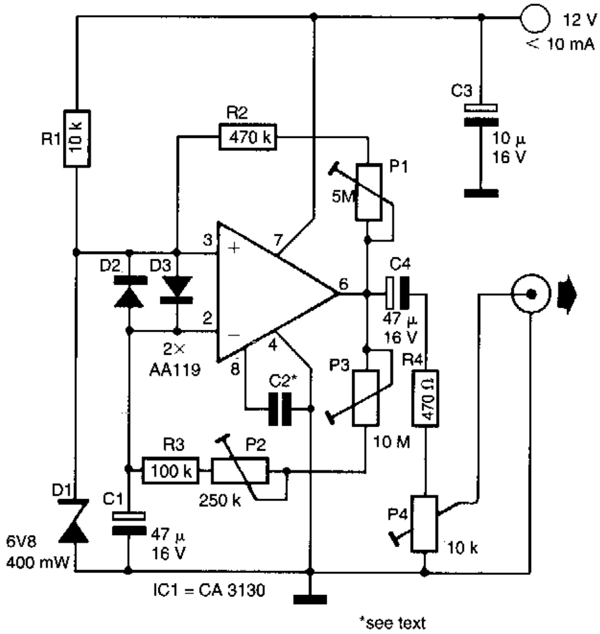
67

Noise-Generator Circuits

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Noise Generator

NOISE GENERATOR



303 CIRCUITS ***FIG. 67-1***

This circuit generates noise pulses that are suitable for test purposes, etc. A zener diode is used as a noise source. IC1 is a relaxation oscillator. P1 determines noise bandwidth, and P2 and P3 the noise amplification. Current consumption is 10 mA @ 12 Vdc.

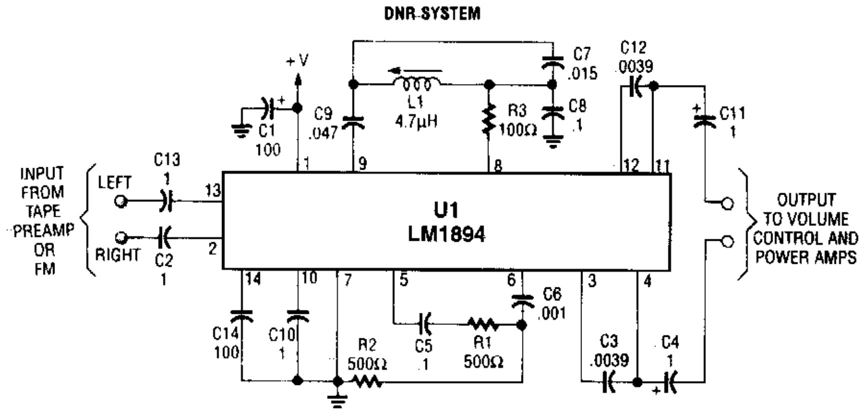
68

Noise-Limiting Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Audio Dynamic Noise-Reduction System
Amplified Noise Limiter for SW Receivers
Receiver AF Noise Limiter for Low-Level Signals
Simple Noise Limiter for Receivers

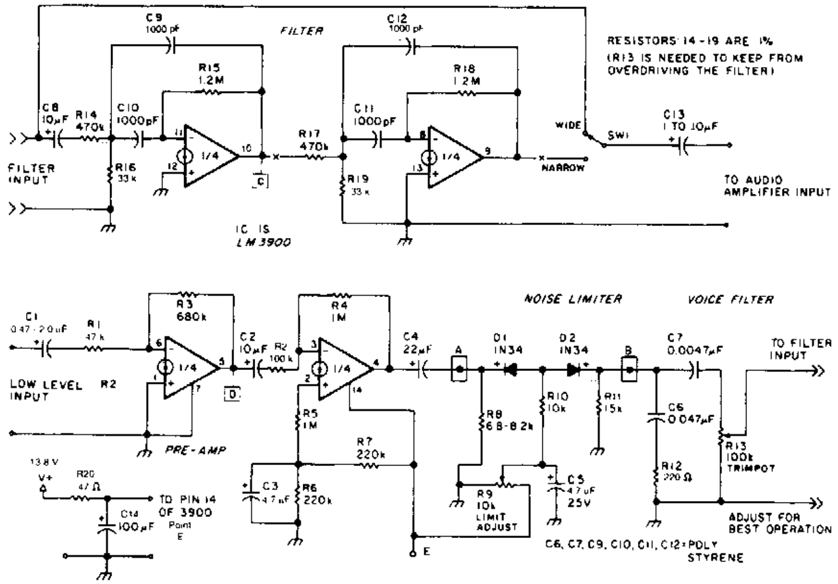
AUDIO DYNAMIC NOISE-REDUCTION SYSTEM



POPULAR ELECTRONICS ***FIG. 68-1***

U1 is a dedicated IC (National Semiconductor) that achieves up to 10 dB noise reduction by an adaptive bandwidth scheme and a psycho acoustic masking technique.

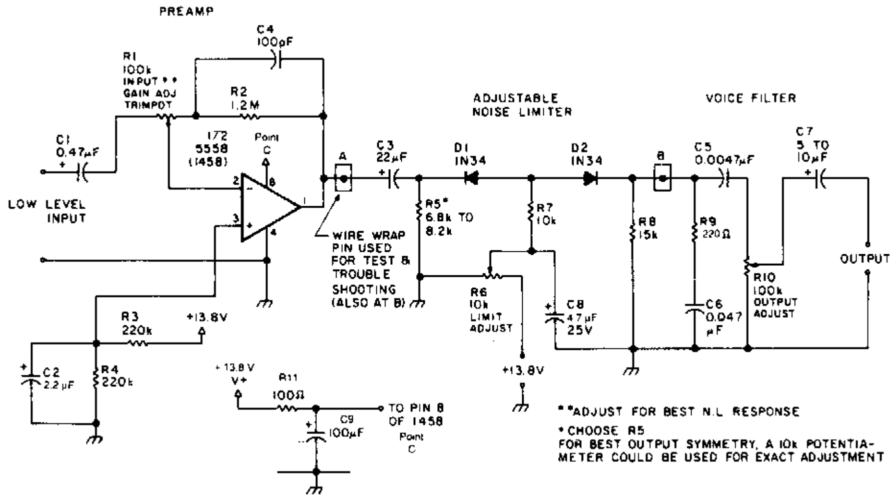
AMPLIFIED NOISE LIMITER FOR SW RECEIVERS



73 AMATEUR RADIO TODAY ***FIG. 68-2***

The noise limiter circuit has a preamplifier clipper, and a switchable audio bandpass filter. Audio levels in the 5- to 50-mV range are amplified in a preamp to several volts p-p, fed to a clipper, voice band filter, then to a narrow band active filter which can be switched in and out of the circuit.

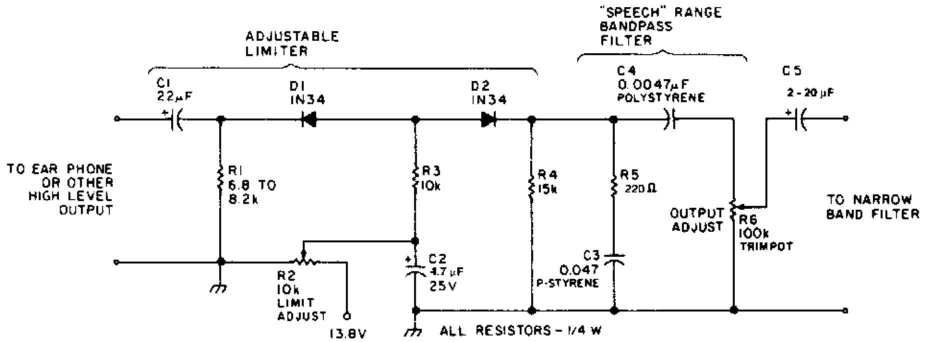
RECEIVER AF NOISE LIMITER FOR LOW-LEVEL SIGNALS



73 AMATEUR RADIO TODAY *FIG. 68-3*

A preamplifier in the audio frequency range amplifies a noisy audio signal to drive a diode clipper. Suitable audio input levels would be in the 10-mV to 1-V range.

SIMPLE NOISE LIMITER FOR RECEIVERS



73 AMATEUR RADIO TODAY *FIG. 68-4*

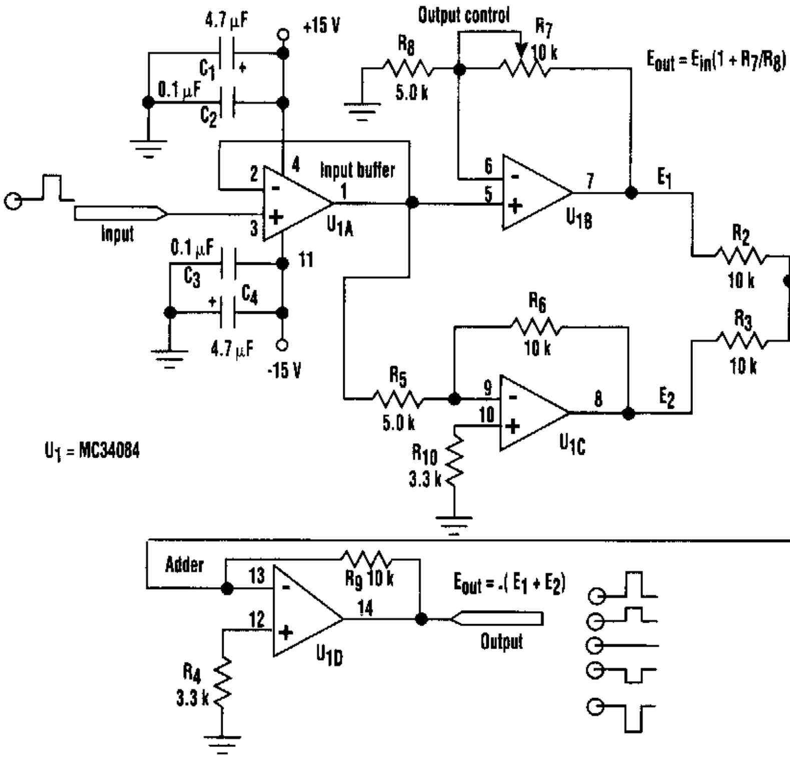
This circuit uses a diode series clipper to limit noise peaks on a received signal. It is best used where several volts p-p of audio signal are available.

69

Operational-Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

POLARITY GAIN ADJUSTMENT



ELECTRONIC DESIGN

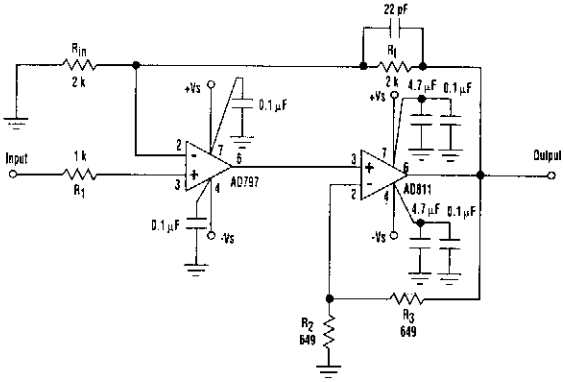
FIG. 69-1

By adjusting one potentiometer, this circuit's output can be varied from a positive-going version of the input signal, smoothly through zero output, then to a negative-going version of the input (see the figure). If the input signal is a positive pulse of, for example, +2-V peak, the output pulse amplitude can be smoothly varied from +2-V through ground (no output) to a –2-V peak.

Taking a closer look at the setup, assume that the signal has a +2-V peak input. The A section of the quad op amp is an input buffer, op amp C provides a fixed negative-going output of –4-V peak, and op amp B supplies a positive-going output that varies from +2-V to +6-V peak. The D section adds the B and C outputs. Thus, by varying the B output, the circuit output varies smoothly from –2-V to +2-V peak.

The circuit can, of course, also be used as a 0°/180° phase switcher. For instance, with a ground-centered sine-wave input of 4V p-p, the output varies from 4-V p-p in phase with the input, smoothly through 0 V, to 4V p-p 180° out of phase with the input.

FAST COMPOSITE AMPLIFIER



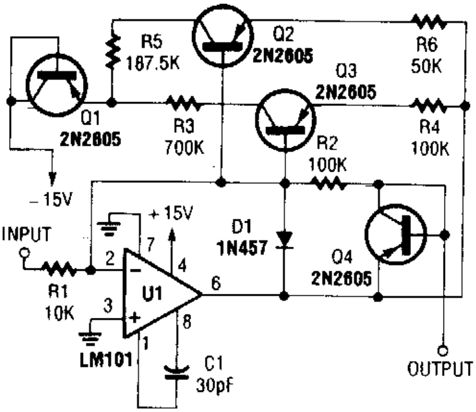
ELECTRONIC DESIGN *FIG. 69-2*

An ultra-low-noise, low-distortion op amp—the AD797—is combined with the AD811 op amp, which offers a high bandwidth and a 100-mA output drive capability. The composite-amplifier circuit serves quite well when driving high resolution ADC's and ATE systems.

The fast AD811 operates at twice the gain of the AD797 so that the slower amplifier need only slew one-half of the total output swing. Using the component values shown, the circuit is capable of better than –90 dB THD with a ±5-V, 500-kHz output signal. If a 100-kHz sine-wave input is used, the circuit will drive a 600-Ω load to a level of 7 V rms with less than –109 dB THD, as well as a 10-kΩ load at less than –117 dB THD.

The device can be modified to supply an overall gain of 5 by changing both the R_f/R_{in} ratio and R_3/R_2 ratio to 4:1. This raises the gains of AD811 and the total circuit while maintaining the AD797 at unity gain. If only the R_f/R_{in} ratio is changed, the circuit might become unstable. In contrast, if only the R_3/R_2 ratio is varied, the AD797 will then operate at gain. Subsequently, the circuit will have a lower overall bandwidth. R_1 should be equal to the parallel combination of R_{in} and R_f.

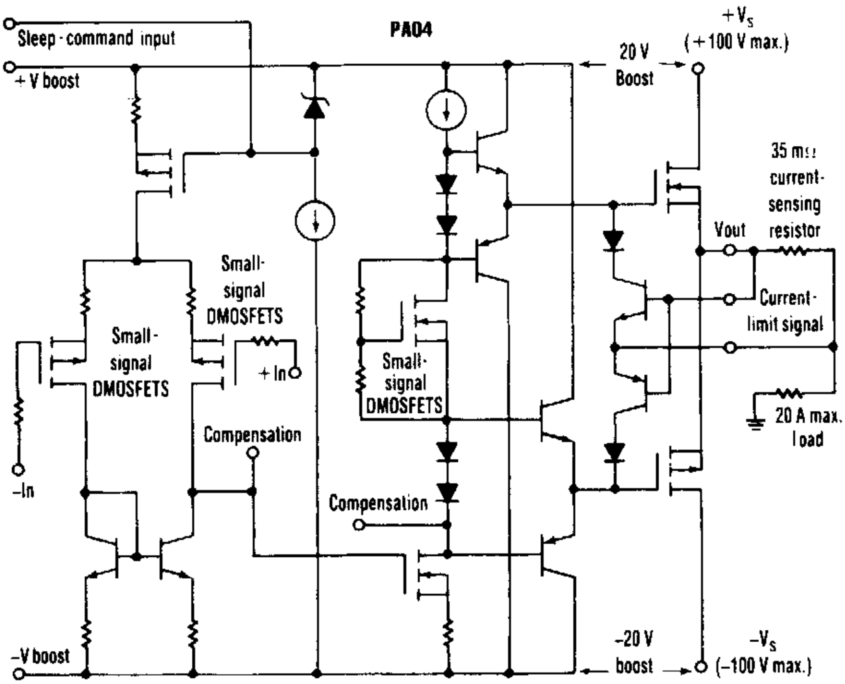
NONLINEAR OPERATIONAL AMPLIFIER WITH TEMPERATURE-COMPENSATED BREAKPOINTS



Using resistor and transistor feedback elements, this operational amplifier circuit can be used as a nonlinear amplifier. R4 and R6 can be varied to change breakpoints, as required.

POPULAR ELECTRONICS *FIG. 69-3*

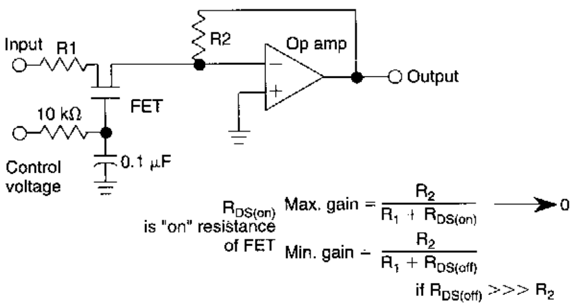
POWER OP AMP



ELECTRONIC DESIGN ***FIG. 69-4***

This circuit from Apex Microtechnology can deliver 180 V p-p @ 90 kHz into a 4-Ω load. The PA04 can deliver 400-W RMS into an 8-Ω load with low THD at frequencies beyond 20 kHz.

VARIABLE GAIN OP-AMP CIRCUIT

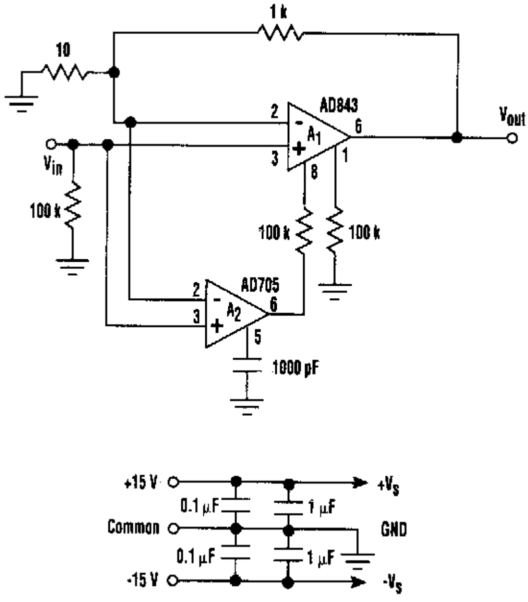


A JFET acts as a variable attenuator for this op amp. Maximum gain is:

$$\frac{R_2}{R_1 + R_{DS(ON)}}$$

ELECTRONICS NOW ***FIG. 69-5***

LOW NOISE AND DRIFT COMPOSITE AMP

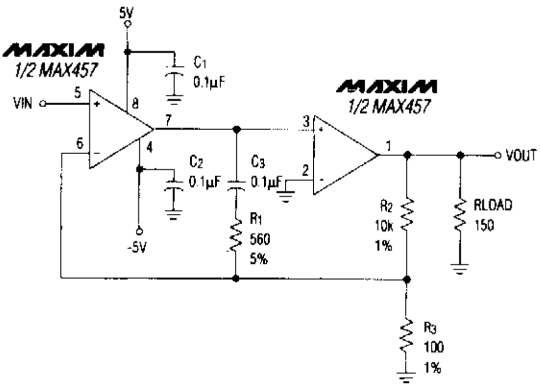


ELECTRONIC DESIGN FIG. 69-6

This circuit offers the best of both worlds. It can be combined with a low input offset voltage and drift without degrading the overall system's dynamic performance. Compared to a standalone FET input operational amplifier, the composite amplifier circuit exhibits a 20-fold improvement in voltage offset and drift.

In this circuit arrangement, A1 is a high-speed FET input op amp with a closed-loop gain of 100 (the source impedance was arbitrarily chosen to be 100 kΩ). A2 is a SuperBeta bipolar input op amp. It has good dc characteristics, biFET-level input bias current, and low noise. A2 monitors the voltage at the input of A1 and injects current to A1's null pins. This forces A1 to have the input properties of a bipolar amplifier while maintaining its bandwidth and low-input-bias-current noise.

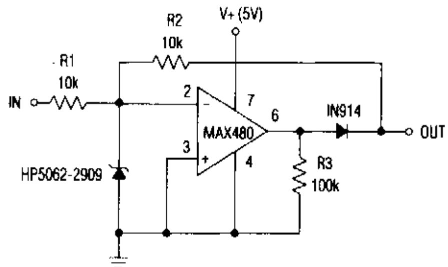
HIGH-GBW OP AMP



MAXIM ENGINEERING JOURNAL FIG. 69-7

You can build a composite amplifier featuring high gain, wide bandwidth, and good dc accuracy by cascading the sections of a dual video amplifier and adding two appropriate phase-compensation components. The op amp drives a 150-Ω load and provides a closed-loop gain of 40 dB.

SINGLE OP-AMP FULL-WAVE RECTIFIER



MAXIM ENGINEERING JOURNAL FIG. 69-8

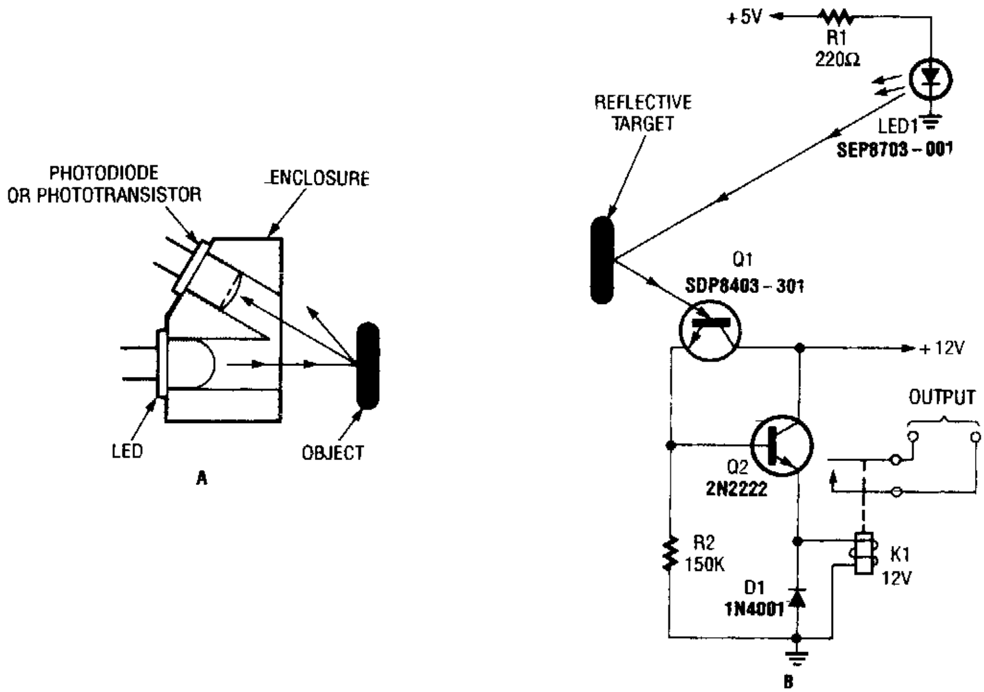
This circuit operates from +5 V and uses a single op amp to deliver a full-wave rectified output of the input signal.

70

Optical Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

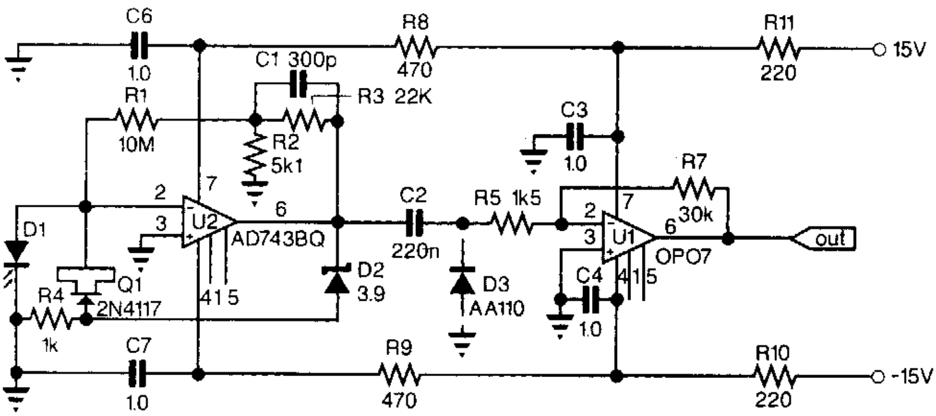
OPTICAL PROXIMITY DETECTOR



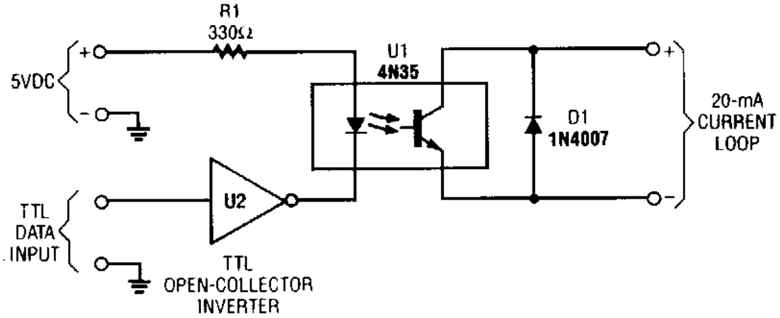
POPULAR ELECTRONICS **FIG. 70-1**

A "reflector" isolator (A) detects the presence of an object by bouncing light off of it. This technique is useful in circuits that detect when an object is close enough to the sensor (B).

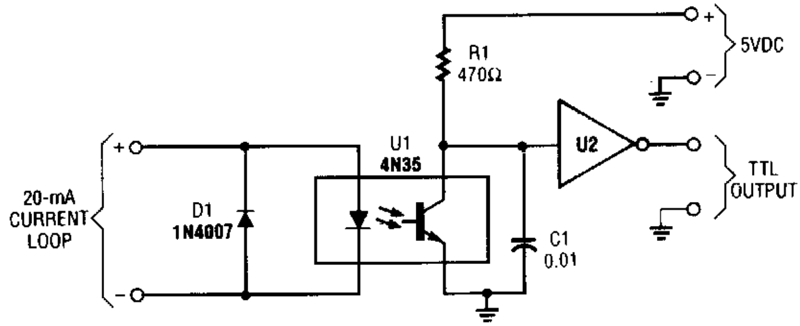
PHOTORECEIVER OPTIMIZED FOR NOISE AND RESPONSE



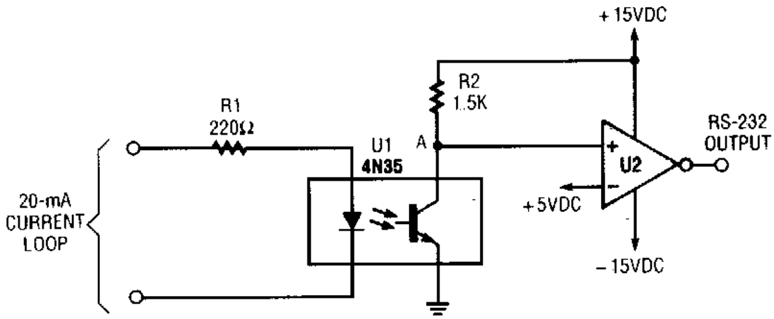
ELECTRONIC ENGINEERING **FIG. 70-2**



A



B

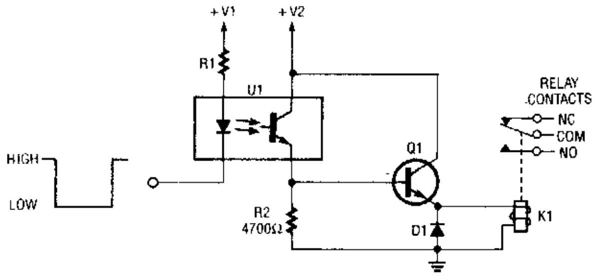


C

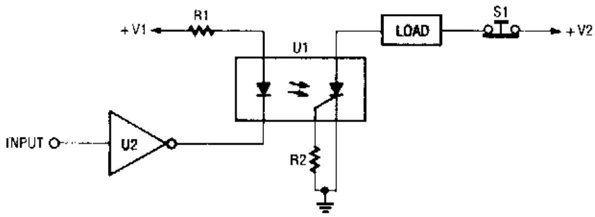
Interfacing equipment, whether TTL, RS-232C, or 20=mA current-loop based, with optoisolators.

POPULAR ELECTRONICS

FIG. 70-3



A

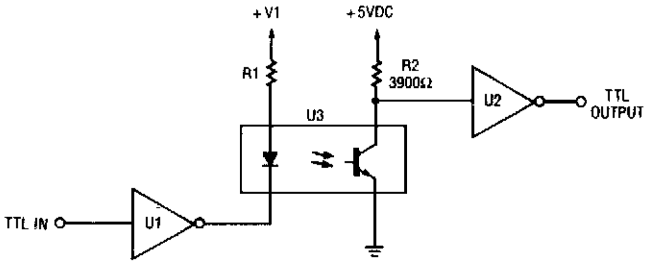


B

Very heavy loads, which can't be powered directly by an optoisolator, might require the use of a relay as shown in A. You can sometimes get away with using a circuit like that shown in B, but it won't turn itself off.

A circuit for isolating a variable resistor is shown. An optoisolator that has an LED and a photoconductive cell (or photoresistor) is used. The current through the LED controls its brightness, which in turn determines the resistance between terminals A and B. The LED current is set by the voltage of the dc power supply and the value of the two resistors (R1 and R2). The fixed resistor (R1) is used to limit the current to a maximum of 20 mA (when the resistance of the potentiometer, R_2, is set to zero ohms), otherwise, the LED might burn out.

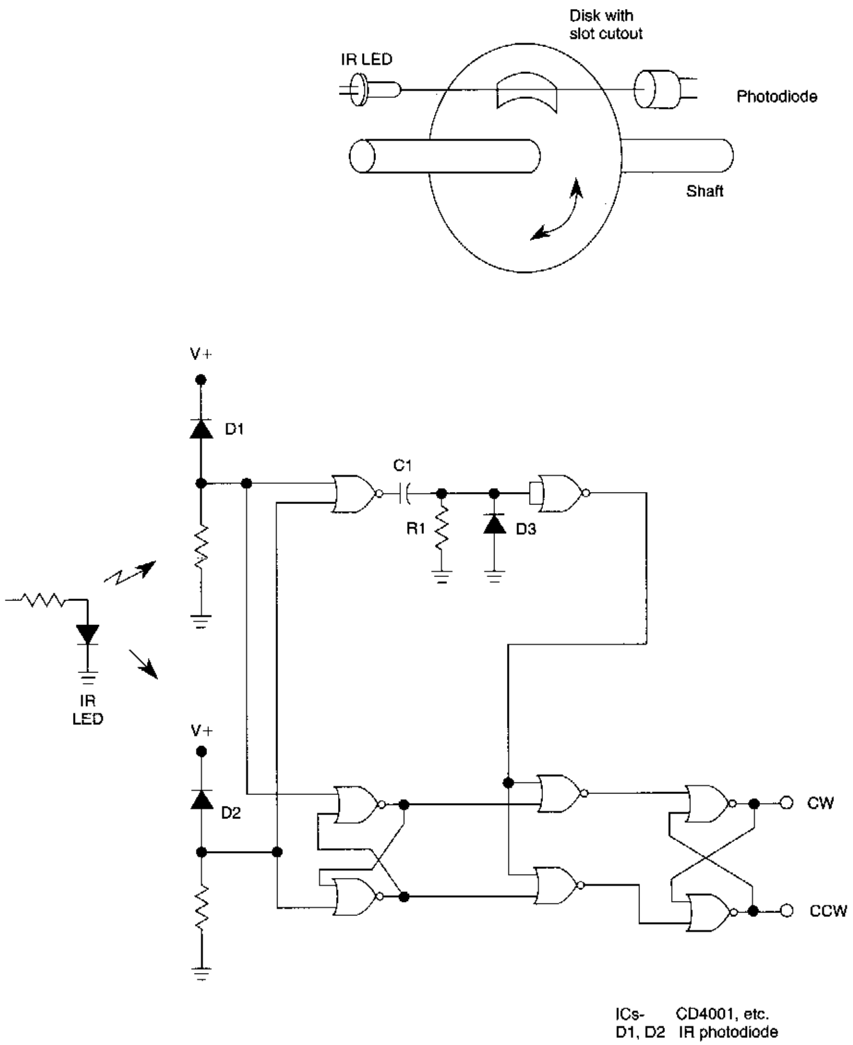
OPTOCOUPLER CIRCUITS



POPULAR ELECTRONICS ***FIG. 70-4***

This circuit is a TTL-to-TTL isolator circuit. The driver circuit is an open-collector TTL inverter (U1). When the input is high, then the output of the inverter is low. Thus, when the input is high, the output of U1 grounds the cathode end of the LED and causes the LED to turn on.

OPTICAL DIRECTION DISCRIMINATOR

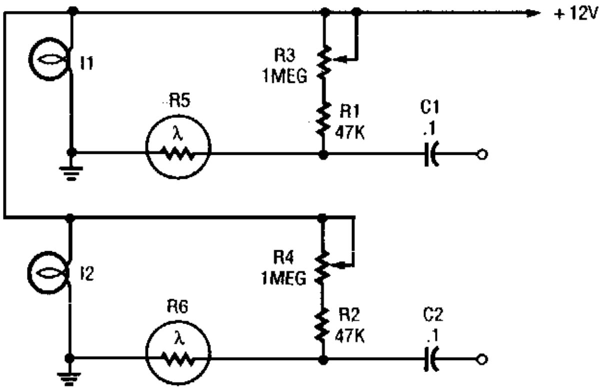


WILLIAM SHEETS ***FIG. 70-5***

The very simple circuit uses only two CD4001 packages, i.e., eight NOR gates and operates in the following way: Pulse streams are fed to an RS flip flop generating an output waveform which has a small or large duty cycle depending on the direction of rotation. The same input pulses are also fed to a NOR gate, which "adds" the two pulse trains.

The rising edges of this waveform are used to produce short positive pulses from the circuit consisting of R1, C1, D3, and a NOR gate used as an inverter. This is used to "sample" the outputs of the flip flop to detect the direction of rotation. The output, whose duty cycle is large, forces the sampling NOR gate to generate a pulse train which sets (or resets) the second RS flip-flop continuously giving a permanent indication of the direction of rotation.

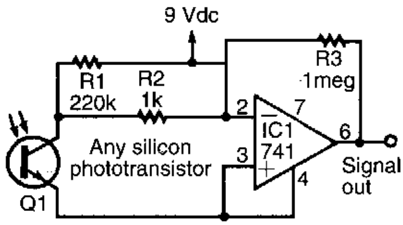
OPTICAL SAFETY CIRCUIT SWITCHES



POPULAR ELECTRONICS ***FIG. 70-6***

Use of two LDR devices replaces the two pushbuttons used in safety switches. The lamps provide light sources for the LDR devices.

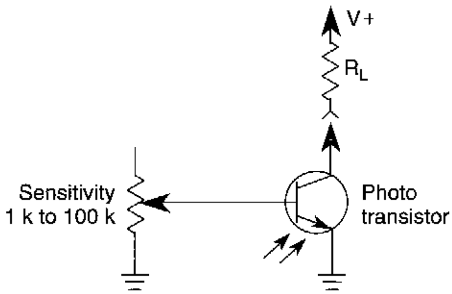
SIMPLE AMPLIFIER FOR PHOTOTRANSISTORS



ELECTRONICS NOW ***FIG. 70-7***

This simple amplifier will work well with just about any phototransistor. The 741, although designed to operate with a split supply, will work with a single-sided supply as well.

VARIABLE-SENSITIVITY PHOTOTRANSISTOR CIRCUIT



ELECTRONICS NOW ***FIG. 70-8***

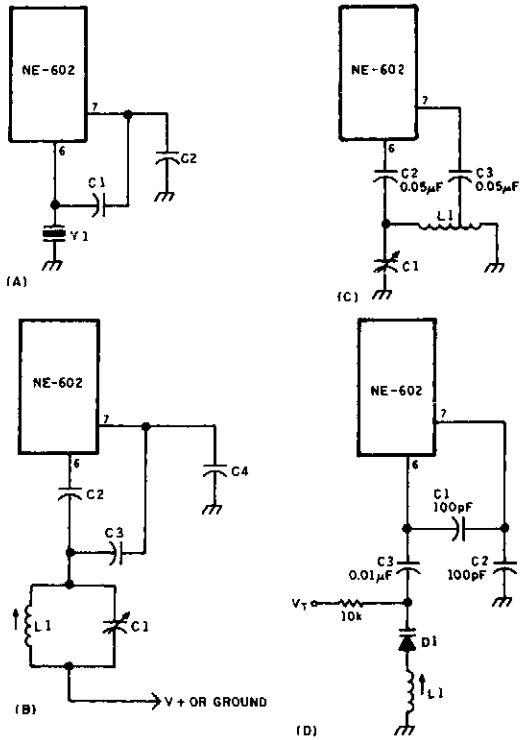
A variable resistor is used to vary the light-level response of a phototransistor. Phototransistors are more light sensitive than photodiodes, but they generally have poorer frequency response.

71

Oscillator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

NE602 LOCAL OSCILLATOR CIRCUITS

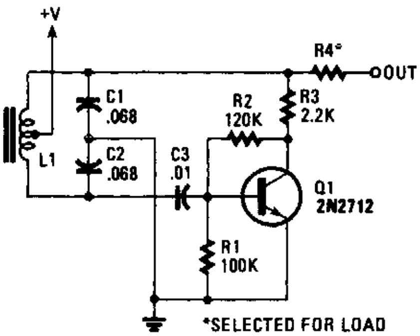


73 AMATEUR RADIO TODAY

FIG. 71-1

Local oscillator circuits for the NE602.

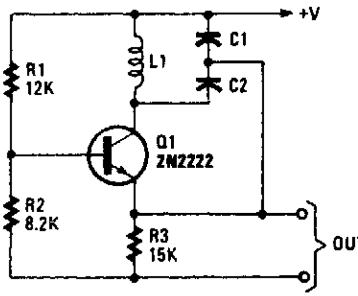
LC AUDIO OSCILLATOR



POPULAR ELECTRONICS

FIG. 71-2

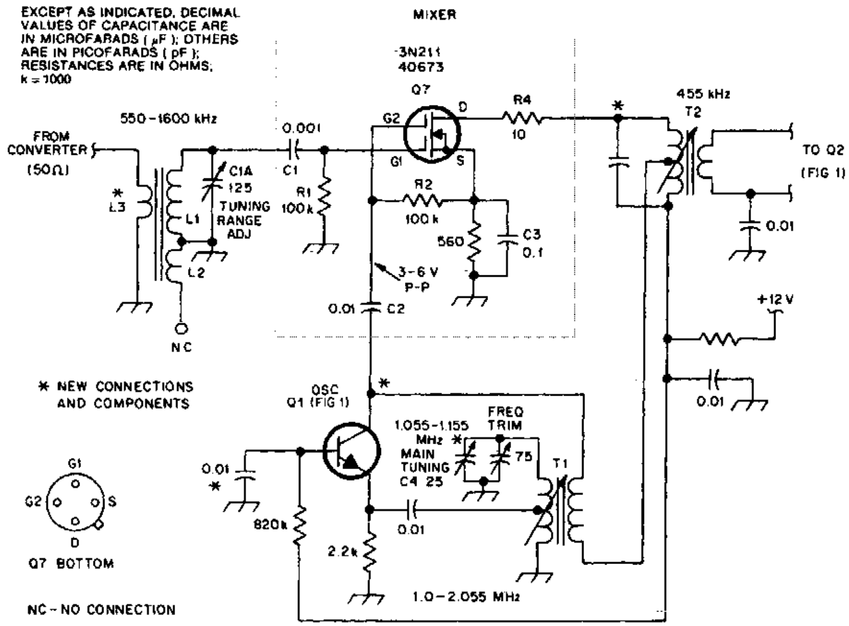
COLPITTS OSCILLATOR



POPULAR ELECTRONICS

FIG. 71-3

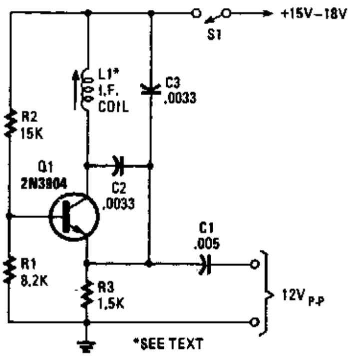
MOSFET MIXER-OSCILLATOR CIRCUIT FOR AM RECEIVERS



QST ***FIG. 71-4***

This circuit is an improved front end for upgrading a transistor AM receiver. This front end is useful when the radio is to be used as a tuneable IF amplifier with shortwave converters.

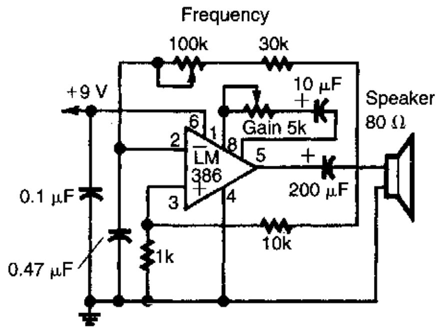
SIMPLE RF TEST OSCILLATOR



POPULAR ELECTRONICS ***FIG. 71-5***

A simple oscillator for IF alignment (455 kHz) can prove useful in field testing or where a standard signal generator is available. L1 should resonate at the desired output frequency with the series combination of C2 and C3.

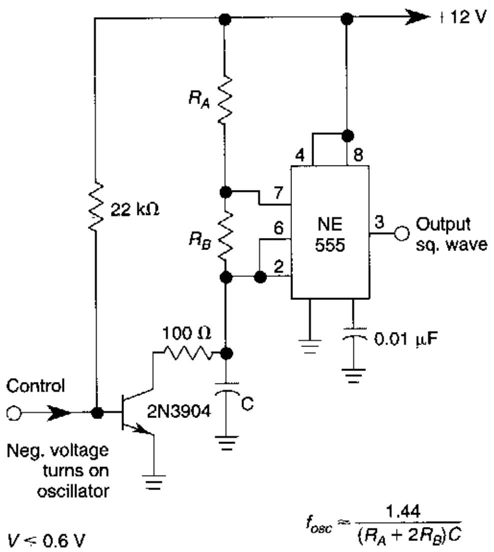
AF POWER OSCILLATOR



RADIO ELECTRONICS ***FIG. 71-6***

An LM386 audio power IC is set up as a feedback oscillator. Any supply from 6 to 12 V can be used. The circuit can drive a loudspeaker.

GATED 1-kHz OSCILLATOR (NORMALLY OFF)

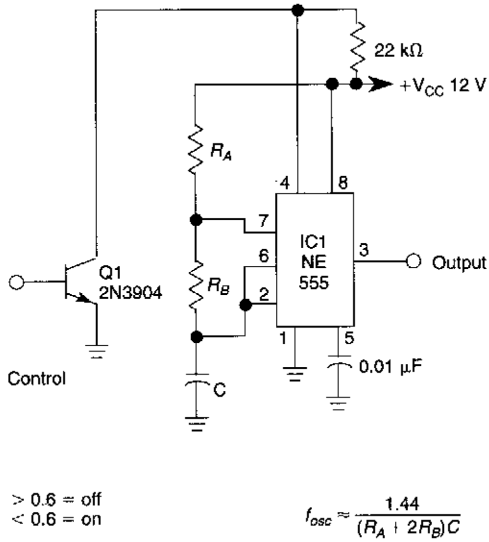


$$f_{osc} \approx \frac{1.44}{(R_A + 2R_B)C}$$

ELECTRONICS NOW ***FIG. 71-7***

This gated 1-kHz oscillator offers "press-to-turn-on" operation, A, and waveforms at the output of pin 3 and across C1, B.

GATED 1-kHz OSCILLATOR (NORMALLY ON)

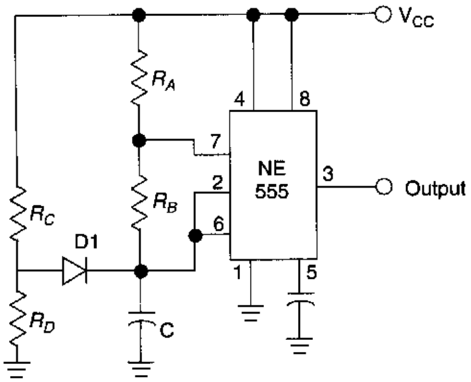


$$f_{osc} \approx \frac{1.44}{(R_A + 2R_B)C}$$

ELECTRONICS NOW ***FIG. 71-8***

This gated 1-kHz oscillator offers "press-to-turn-off" operation, A, and waveforms at the output of pin 3 and across C1, B.

PRECISION LF OSCILLATOR



$$\frac{R_D}{R_C + R_D} = \left(\frac{V_{CC}}{3} + 0.6\right)$$

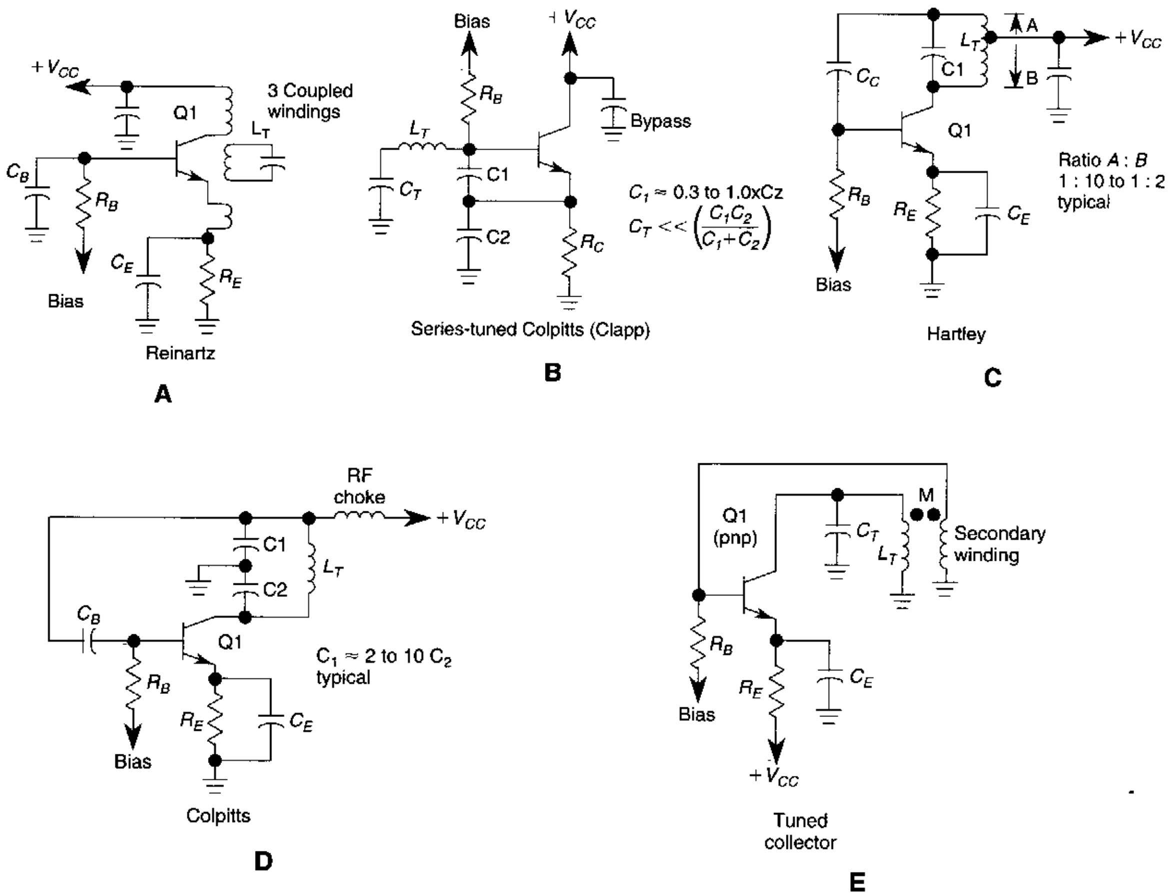
$$R_C + R_D << R_A + R_B$$

$$T \approx \frac{1.44}{(R_A + 2R_B)C}$$

ELECTRONICS NOW ***FIG. 71-9***

Using R1, R7, and D1 to preset C1 to one third of the supply voltage, this circuit avoids a longer first cycle period than subsequent cycles.

BASIC OSCILLATOR CIRCUITS



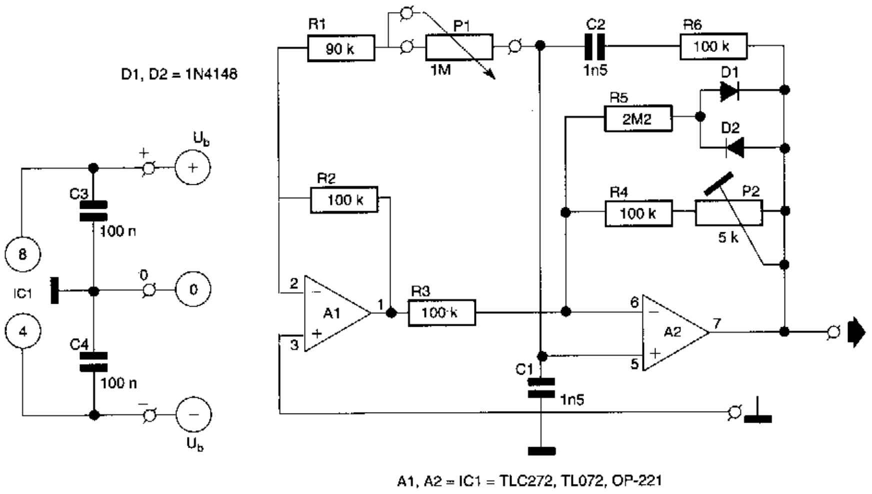
ELECTRONICS NOW **FIG. 71-10**

Five basic types of LC oscillators are shown. The frequency can be changed by using the formula:

$$f = \frac{1}{2\pi L_{\text{effective}} C_{\text{effective}}}$$

where $L_{\text{effective}}$ = equivalent inductance
$C_{\text{effective}}$ = equivalent capacitance

VARIABLE WIEN-BRIDGE OSCILLATOR

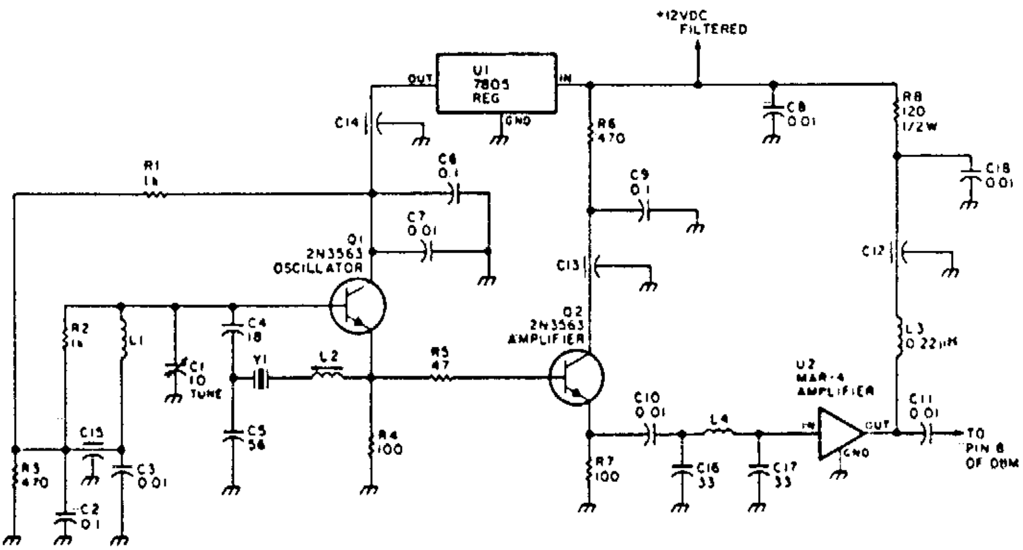


303 CIRCUITS

FIG. 71-11

This circuit uses a single potentiometer to tune a 300- to 3000-Hz range. A FET op amp is used at A1 and A2. The upper frequency limit is determined by the gain-bandwidth product of the op amps.

LOCAL OSCILLATOR FOR DOUBLE BALANCED MIXERS

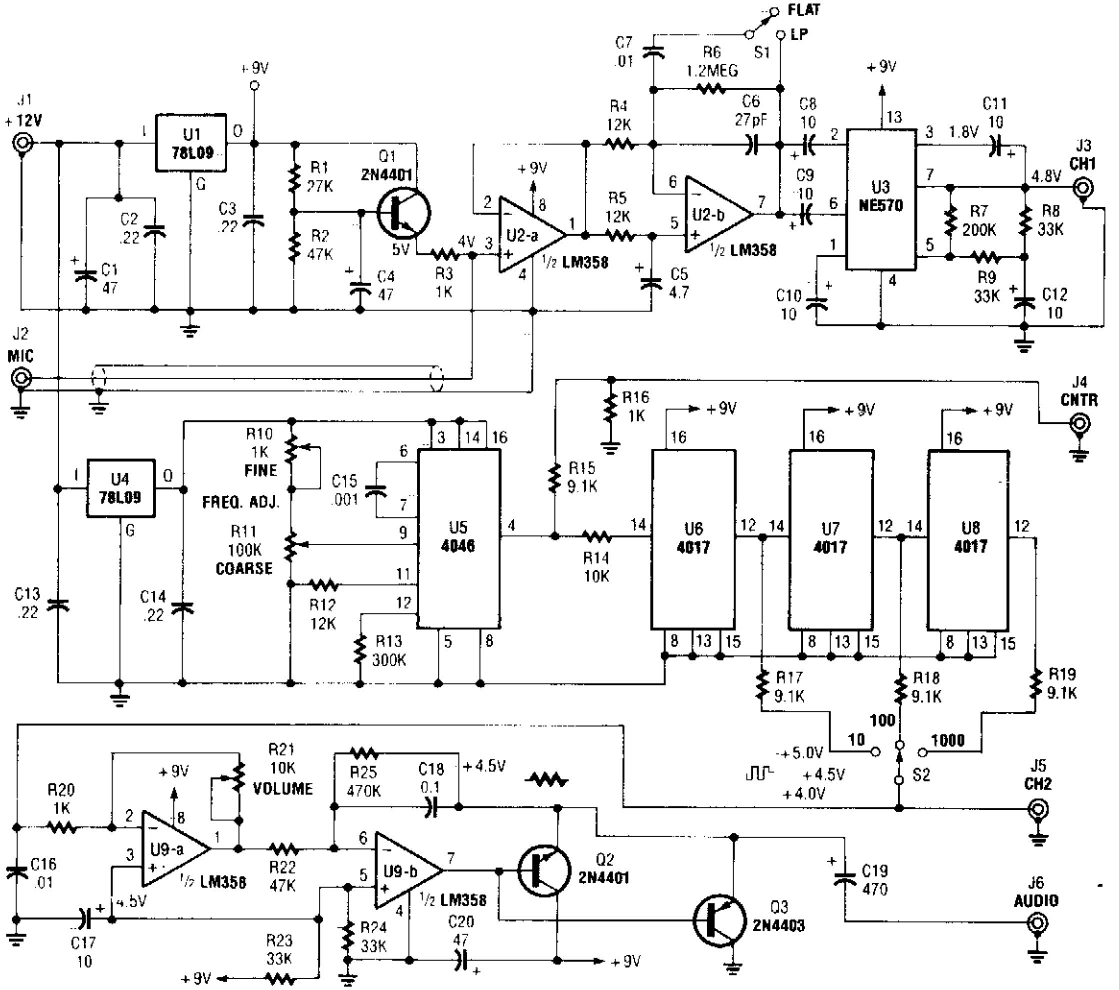


73 AMATEUR RADIO TODAY

FIG. 71-12

This circuit has an amplifier to supply +10 dBm to an SBL series (Mini-circuits) or similar type doubly-balanced mixer assembly. This circuit has values shown for ≈80- to 90-MHz crystals, although values of oscillator circuit constants can be scaled for higher or lower frequencies.

PRECISION AUDIO-FREQUENCY GENERATOR



POPULAR ELECTRONICS *FIG. 71-13A*

The precision audio-frequency generator consists of several subcircuits—an audio-amplifier/filter circuit, an automatic level control, a variable voltage-controlled oscillator, a frequency divider circuit, an integrator, and an audio output amplifier.

An electret microphone element is used to pick up the audio tone produced by the instrument. That signal is then fed to an amplifier/filter/level-controlled circuit and output via channel 1 (CH1) to an oscilloscope for display.

The variable voltage-controlled oscillator (VCO) is used to produce a signal of from less than 10 kHz to more than 99 kHz. The VCO output is fed to a digital frequency counter for display, and is also routed to a chain of frequency dividers, where the signal is divided by 10, 100, or 1,000, depending on the setting of a selector switch.

Note/ Octave	Key#	Hertz	Stretch in Cents	Note/ Octave	Key#	Hertz	Stretch in Cents
A/0	1	27.184	−20	F/4	45	349.03	− 1
B♭/0	2	28.817	−19	G♭/4	46	369.78	− 1
B/0	3	30.548	−18	G/4	47	391.77	− 1
C/1	4	32.384	−17	A♭/4	48	415.07	− 1
D♭/1	5	34.329	−16	A/4	49	440.00	0
D/1	6	36.391	−15	B♭/4	50	466.16	0
E♭/1	7	38.578	−14	B/4	51	493.88	0
E/1	8	40.895	−13	C/5	52	523.25	0
F/1	9	43.352	−12	D♭/5	53	554.37	0
G♭/1	10	45.956	−11	D/5	54	587.33	0
G/1	11	48.717	−10	E♭/5	55	622.61	+ 1
A♭/1	12	51.644	− 9	E/5	56	659.64	+ 1
A/1	13	54.746	− 8	F/5	57	698.86	+ 1
B♭/1	14	58.035	− 7	G♭/5	58	740.42	+ 1
B/1	15	61.522	− 6	G/5	59	784.44	+ 1
C/2	16	65.180	− 6	A♭/5	60	831.57	+ 2
D♭/2	17	69.096	− 5	A/5	61	881.02	+ 2
D/2	18	73.204	− 5	B♭/5	62	933.41	+ 2
E♭/2	19	77.602	− 4	B/5	63	988.91	+ 2
E/2	20	82.217	− 4	C/6	64	1047.7	+ 2
F/2	21	87.106	− 4	D♭/6	65	1110.7	+ 3
G♭/2	22	92.285	− 4	D/6	66	1176.7	+ 3
G/2	23	97.773	− 4	E♭/6	67	1246.7	+ 3
A♭/2	24	103.65	− 3	E/6	68	1321.6	+ 4
A/2	25	109.81	− 3	F/6	69	1400.1	+ 4
B♭/2	26	116.34	− 3	G♭/6	70	1484.3	+ 5
B/2	27	123.26	− 3	G/6	71	1572.5	+ 5
C/3	28	130.59	− 3	A♭/6	72	1667.0	+ 6
D♭/3	29	138.35	− 3	A/6	73	1766.1	+ 6
D/3	30	146.58	− 3	B♭/6	74	1872.2	+ 7
E♭/3	31	155.29	− 3	B/6	75	1984.7	+ 8
E/3	32	164.53	− 3	C/7	76	2103.9	+ 9
F/3	33	174.31	− 3	D♭/7	77	2230.3	+10
G♭/3	34	184.73	− 2.5	D/7	78	2230.2	+10
G/3	35	195.71	− 2.5	E♭/7	79	2506.3	+12
A♭/3	36	207.41	− 2	E/7	80	2656.9	+13
A/3	37	219.75	− 2	F/7	81	2818.1	+15
B♭/3	38	232.81	− 2	G♭/7	82	2989.2	+17
B/3	39	246.66	− 2	G/7	83	3170.6	+19
C/4	40	261.32	− 2	A♭/7	84	3363.0	+21
D♭/4	41	276.86	− 2	A/7	85	3567.1	+23
D/4	42	293.33	− 2	B♭/7	86	3783.6	+25
E♭/4	43	310.86	− 1.5	B/7	87	4013.2	+27
E/4	44	329.44	− 1	C/8	88	4259.2	+30

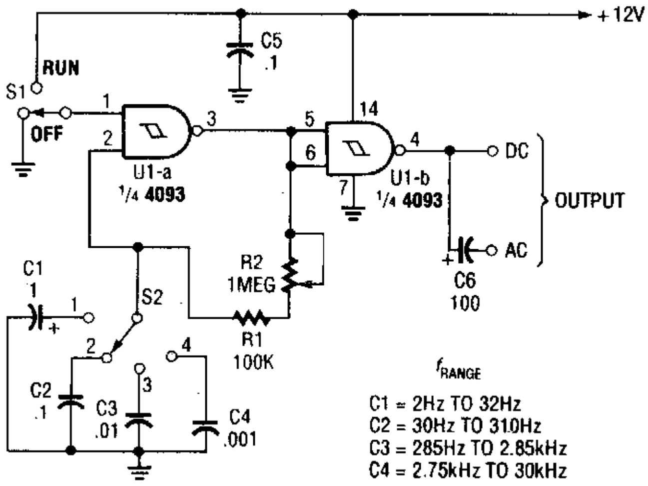
•Standard pitch, A49= 440 Hz
Values shown are stretched for the average piano

POPULAR ELECTRONICS **_FIG. 71-13B_**

From there, the selected signal frequency divides along two paths; one going to CH2 (which feeds the oscilloscope's sweep synchronization input) and to an integrator that converts the square-wave output of the divider into a triangular waveform. The output of the integrator is then amplified and fed to a set of stereo headphones via an audio output jack.

One section of the precision audio-frequency generator uses an electret microphone element to pick up audio from the piano. That signal is then processed and sent to one channel of a dual-trace oscilloscope. The other section of the circuit is used to produce a variable-frequency signal that is fed to a digital frequency counter and, after conditioning, is presented to the second channel of the scope and output to a set of stereo headphones.

CMOS VFO

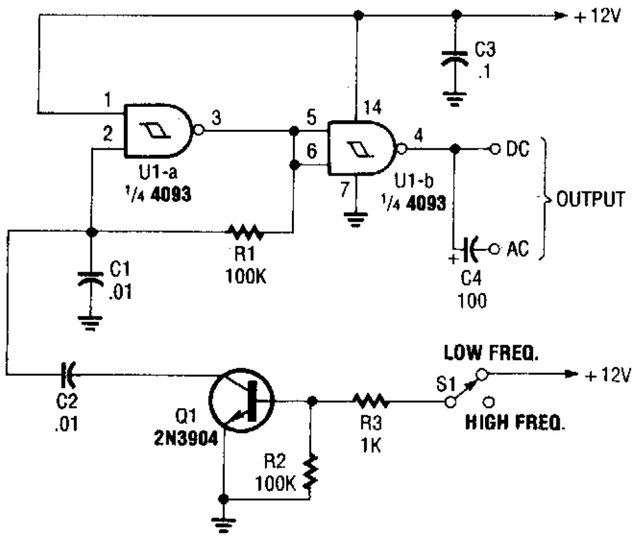


POPULAR ELECTRONICS

FIG. 71-14

The circuit shown has a frequency range of 2 Hz to 30 kHz. R2 is a linear or log potentiometer.

FREQUENCY SWITCHER

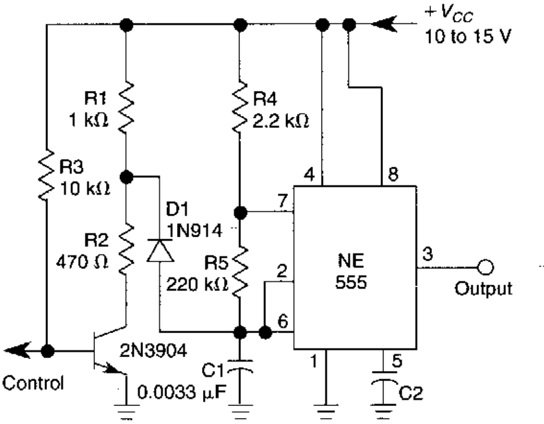


POPULAR ELECTRONICS

FIG. 71-15

This transistor can achieve frequency switching in this CMOS astable oscillator.

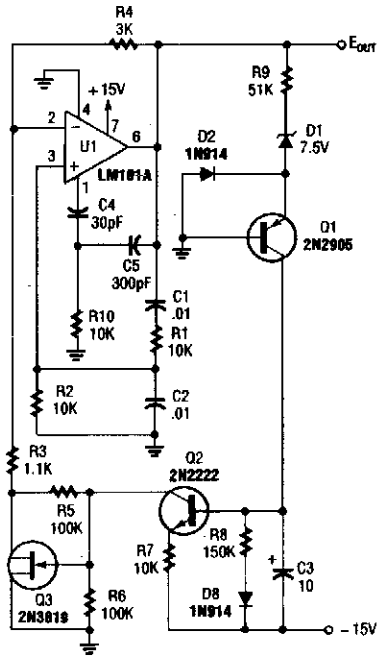
PRECISION GATED OSCILLATOR



ELECTRONICS NOW ***FIG. 71-16***

A 1-kHz gated oscillator with no long "turn-on" cycle is shown. R2, R3, and D1 preset the voltage on tuning capacitor C1 to ⅓ of the supply voltage.

WIEN-BRIDGE AUDIO OSCILLATOR

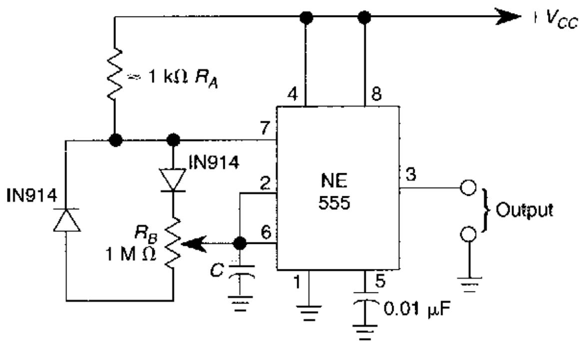


R1 = R2 C1 = C2 $f = \frac{1}{2\pi R1C1}$

POPULAR ELECTRONICS ***FIG. 71-17***

For variable-frequency operation, R1 and R2 can be replaced by a dual potentiometer.

VARIABLE DUTY-CYCLE OSCILLATOR



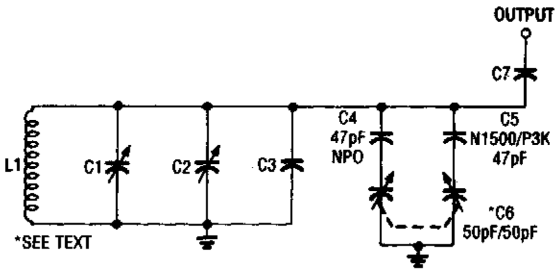
$$T \approx \frac{1.44}{(R_A + 2R_B)C}$$

NOTE: Diodes have the effect of slightly reducing the observed frequency—especially if V_{CC} < 10 V as a result of 0.6 V offset.

ELECTRONICS NOW ***FIG. 71-18***

Using a potentiometer and steering diodes, this 1.2-kHz oscillator will provide 1 to 99% duty cycle. Vary C1 to change frequency.

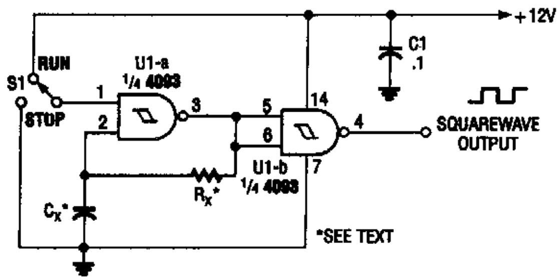
ADJUSTABLE VFO TEMPERATURE COMPENSATOR



POPULAR ELECTRONICS *FIG. 71-19*

Use of a differential capacitor allows temperature compensation of LC circuit using an NPO and N1500 ceramic. C6 is a differential capacitor that has two stators and one common rotor. When one capacitance (stator) is maximum, the other is minimum. L1, C1, C2, and C3 are tuning, trimming, and fixed capacitors, respectively.

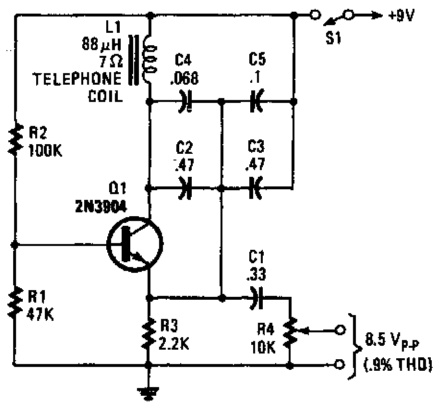
4093 CMOS ASTABLE OSCILLATOR



Two gates of the Quad 4093 are used to make an oscillator. R_x can be from about 5 kΩ to around 10 MΩ. C_x can be from about 10 pF to many µF, the limit being set by the leakage of the capacitor. Frequency is approximately $2.8/R_xC_x$ (R MΩ, Cmfd).

POPULAR ELECTRONICS *FIG. 71-20*

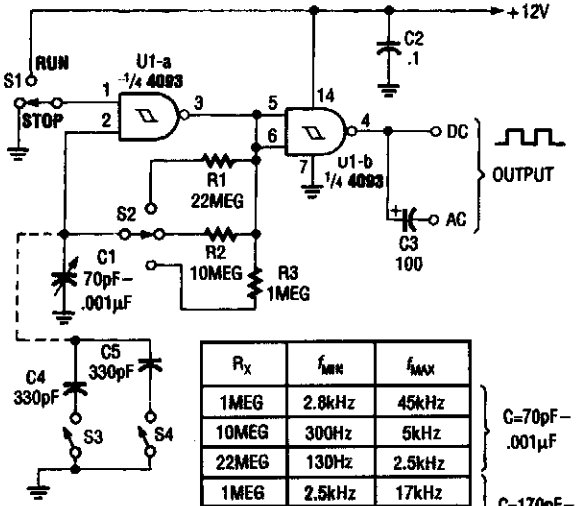
SIMPLE AUDIO TEST OSCILLATOR



An 88-mH surplus telephone toroidal coil is used in a 1-kHz oscillator. Up to 8 V p-p into a high-Z load is available. THD is 0.9%.

POPULAR ELECTRONICS *FIG. 71-21*

4093 CMOS VFO



R_X	f_{MIN}	f_{MAX}	
1MEG	2.8kHz	45kHz	C=70pF–.001µF
10MEG	300Hz	5kHz	C=70pF–.001µF
22MEG	130Hz	2.5kHz	C=70pF–.001µF
1MEG	2.5kHz	17kHz	C=170pF–.0011µF
10MEG	270Hz	1.85kHz	C=170pF–.0011µF
22MEG	115Hz	835Hz	C=170pF–.0011µF

POPULAR ELECTRONICS *FIG. 71-22*

Two gates of a Quad 4093 are used in an astable multivibrator. C1 is a three-gang 365 pF variable capacitor with sections paralleled. S3 and S4 switch in optional extra capacitors.

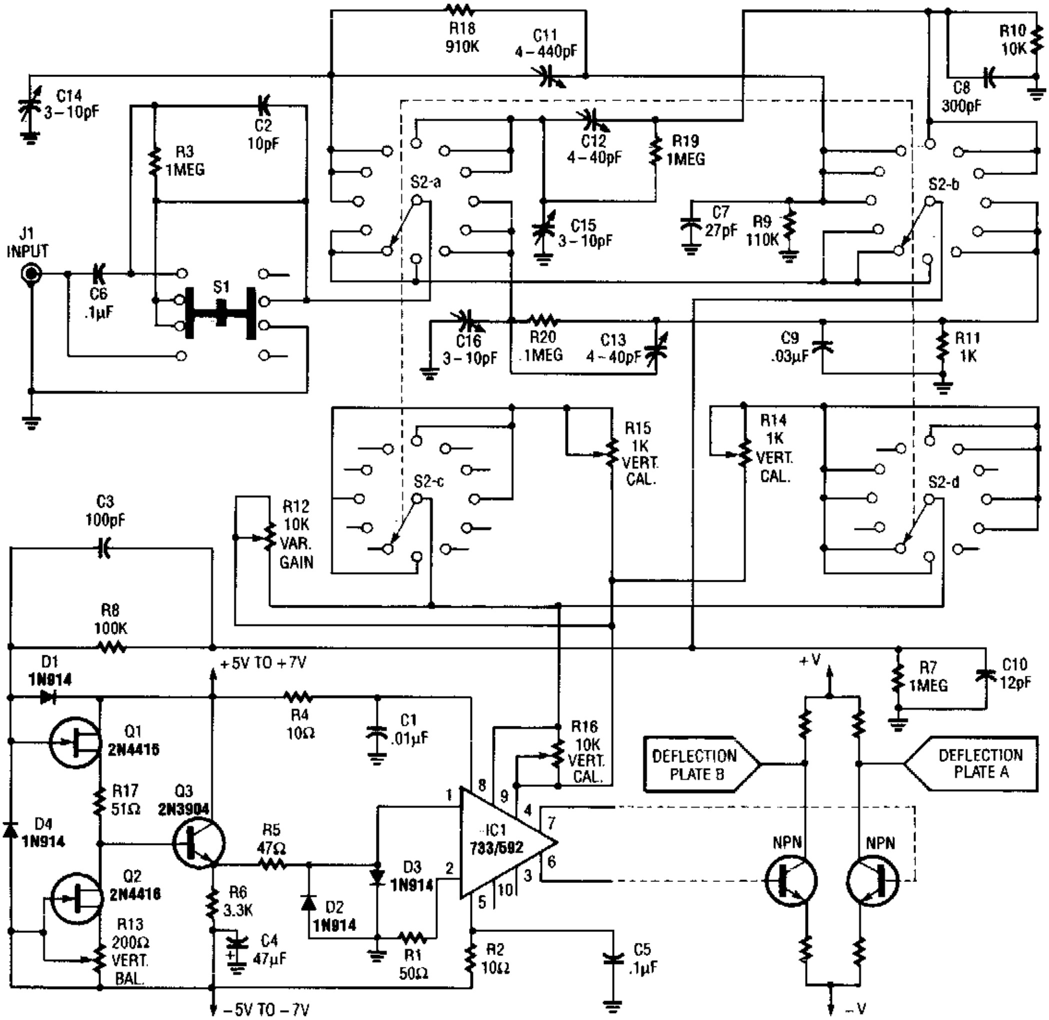
72

Oscilloscope Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Oscilloscope Preamplifier
Simple Spectrum Analyzer Adaptor for Scopes
Simple Oscilloscope Timebase Generator
Trigger Selection Circuit for Oscilloscope Timebase
Variable Gain Amplifier

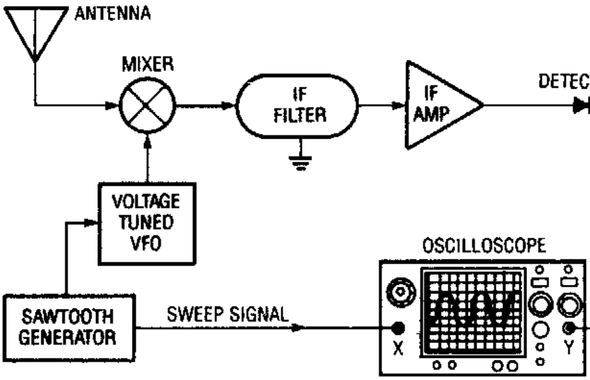
OSCILLOSCOPE PREAMPLIFIER



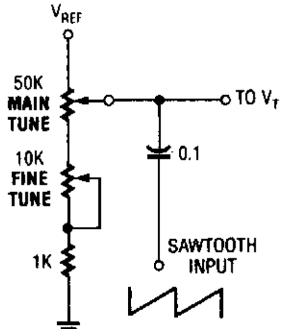
RADIO ELECTRONICS ***FIG. 72-1***

An oscilloscope front-end amplifier can be built with low-cost transistor and video amp ICs. This preamp uses a FET input and compensated attenuators, and has approximately 100-MHz bandwidth, which is adequate for most general-purpose oscilloscopes.

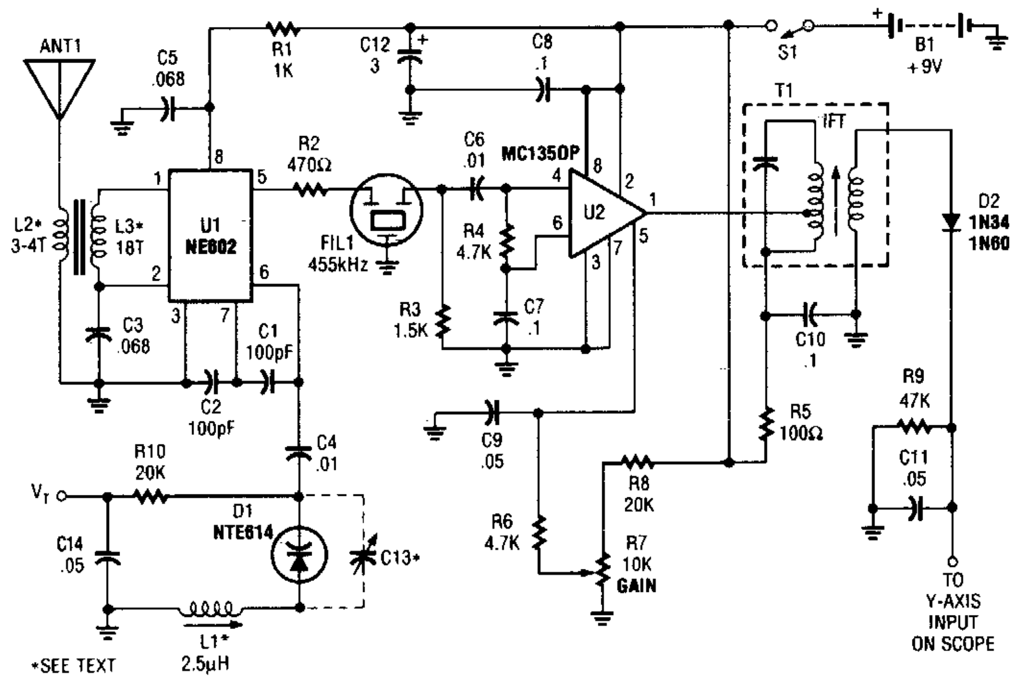
SIMPLE SPECTRUM ANALYZER ADAPTOR FOR SCOPES



Block diagram of a spectrum analyzer.



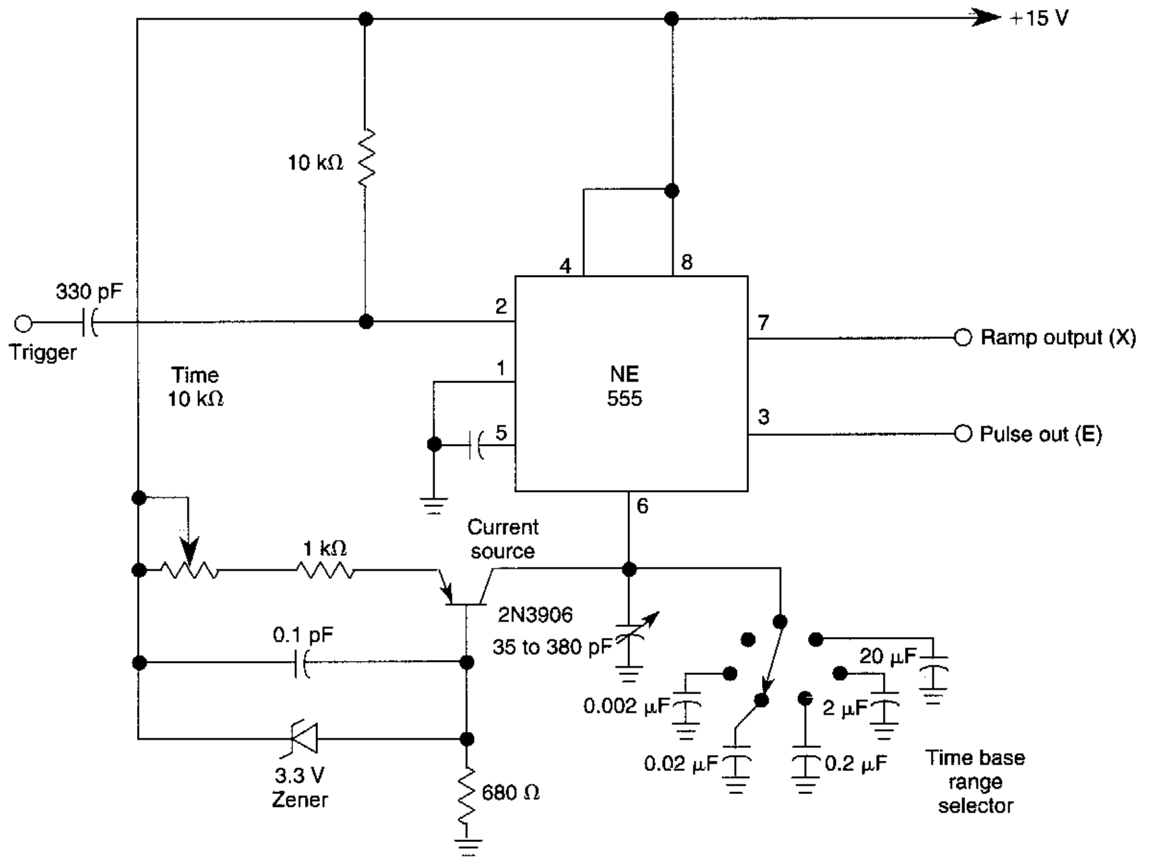
Here's an alternate tuning network for the spectrum analyzer.



POPULAR ELECTRONICS ***FIG. 72-2***

Suitable for monitoring an amateur band or a segment of the radio spectrum, this simple adaptor uses an NE602 mixer-oscillator chip to produce a 455-kHz IF signal, which U2 amplifies, then feeds to detector D2 and the Y axis of an oscilloscope. V_T is used to drive the horizontal axis input of a scope. L2 and L3 are coils suitable for the frequency range in use. For this circuit, coils are shown for the 10- to 15-MHz range. L2 and L3 are wound on Amidon Associates, T-37 or T-50 toroidal cores, and L1 is a commercial or homemade variable inductor, etc.

SIMPLE OSCILLOSCOPE TIMEBASE GENERATOR

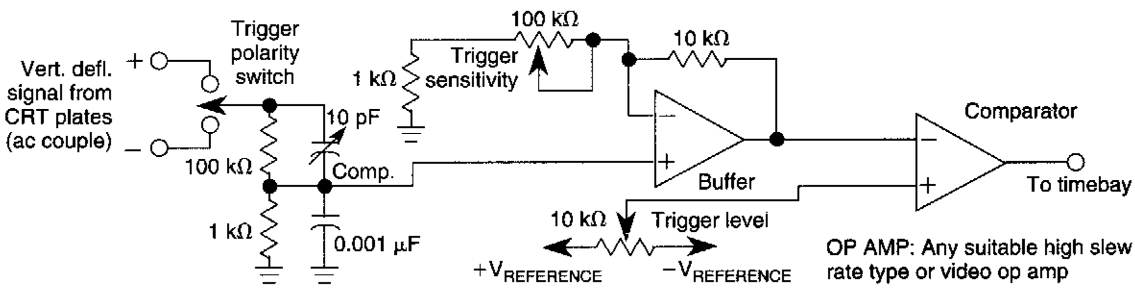


ELECTRONICS NOW

FIG. 72-3

The 555 timer generates both a linear ramp and an output for Z-axis modulations of the CRT electron beam.

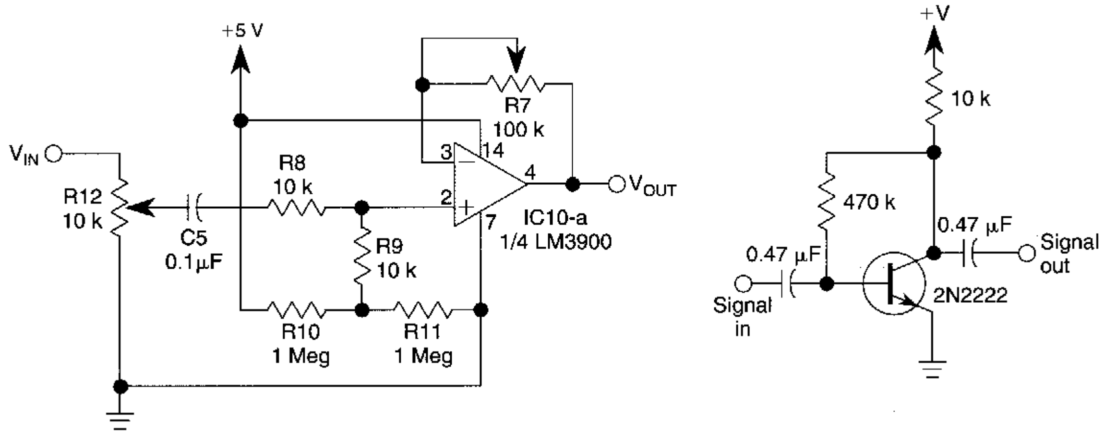
TRIGGER SELECTION CIRCUIT FOR OSCILLOSCOPE TIMEBASE



ELECTRONICS NOW

FIG. 72-4

VARIABLE GAIN AMPLIFIER



ELECTRONICS NOW

FIG. 72-5

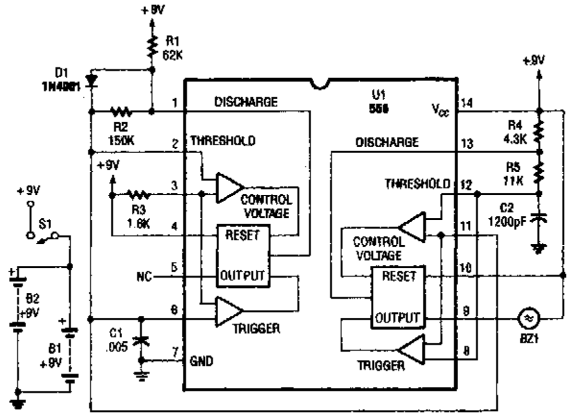
This circuit uses ¼ of an LM3900 to build a simple variable-gain front end for an oscilloscope. R7 is the gain control. Also shown is a simple preamp if you need more than 10X of gain.

73

Pest-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

PEST REPELLER

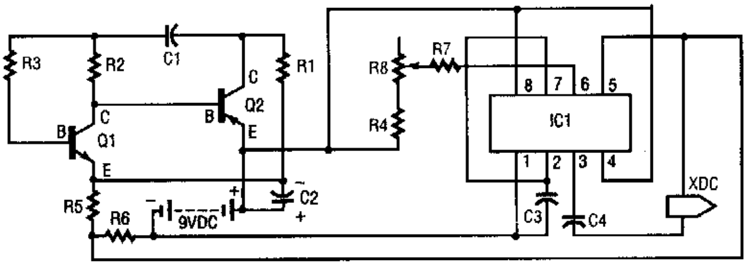


The two timers in the bug repeller have some interesting characteristics. Both of them have their thresholds externally set; the oscillator on the left has a 50% duty cycle and the oscillator on the right acts as a VCO.

1992 PE HOBBYIST HANDBOOK

FIG. 73-1

ULTRASONIC PEST REPELLER



1991 PE HOBBYIST HANDBOOK

FIG. 73-2

This circuit uses two transistors and one IC (555 timer IC) to produce a pulsating ultrasonic frequency. Transistors Q1 and Q2 are connected in a direct-coupled oscillator. The frequency of that oscillator is set by capacitor C1. The oscillator output is taken from the emitter of Q2 to pin 7 of IC1. Transistor Q1 is an npn transistor, and Q2 is a pnp transistor. The signal of pin 7 on IC1 causes the output signal appearing on pin 3 to be modulated or varied by the audio frequency developed by Q1 and Q2. The IC itself is connected as a stable multivibrator with a frequency that is determined by C3. Capacitor C3 sets the basic frequency to be well above the human hearing range (ultrasonic). The combined modulated ultrasonic frequency appears on pin 3 of IC1, where it is coupled by capacitor C4 to the piezoelectric transducer.

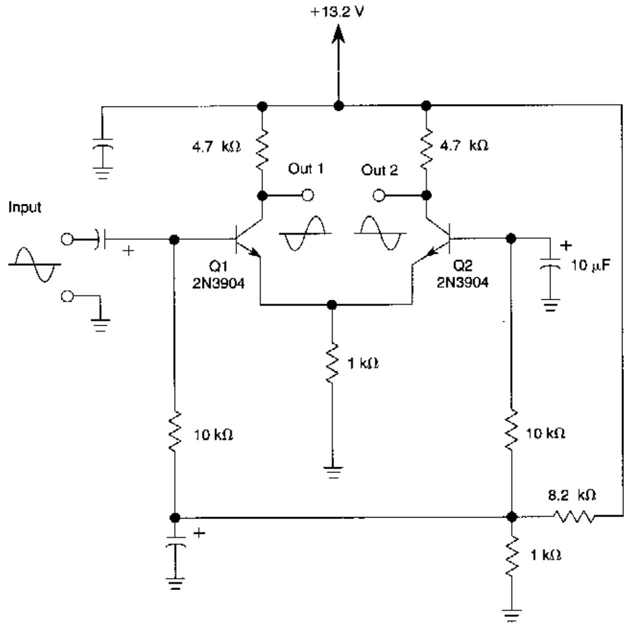
C1, C2	0.1-µF Mylar Capacitor	R2	3.3-MΩ Resistor
C2	1-µF Electrolytic Capacitor	R3, R6	10-kΩ Resistor
C3	0.001-µF Mylar Capacitor	R4, R5	100-Ω Resistor
IC1	555 timer IC	R7	18-kΩ Resistor
Q1	2N3904 Transistor	R8	Potentiometer
Q2	2N3906 Transistor	XDC	Piezoelectric Transducer Disc
R1	4.7-kΩ Resistor	Misc	IC Socket, 9-V Snap, PC Board

74

Phase Shifter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

LONG-TAILED PAIR PHASE-SPLITTER

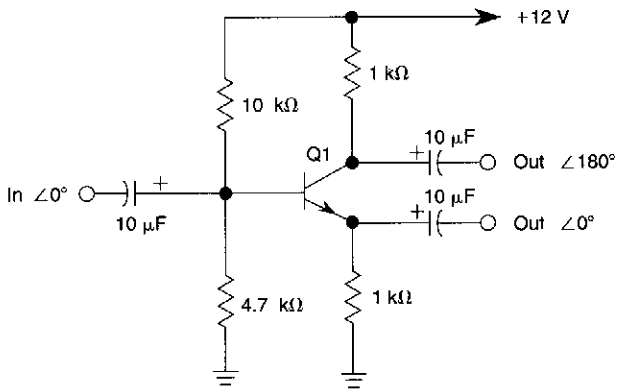


WILLIAM SHEETS

FIG. 74-1

The single-phase input produces out-of-phase outputs at the collectors of Q1 and Q2.

PHASE-SPLITTER CIRCUIT



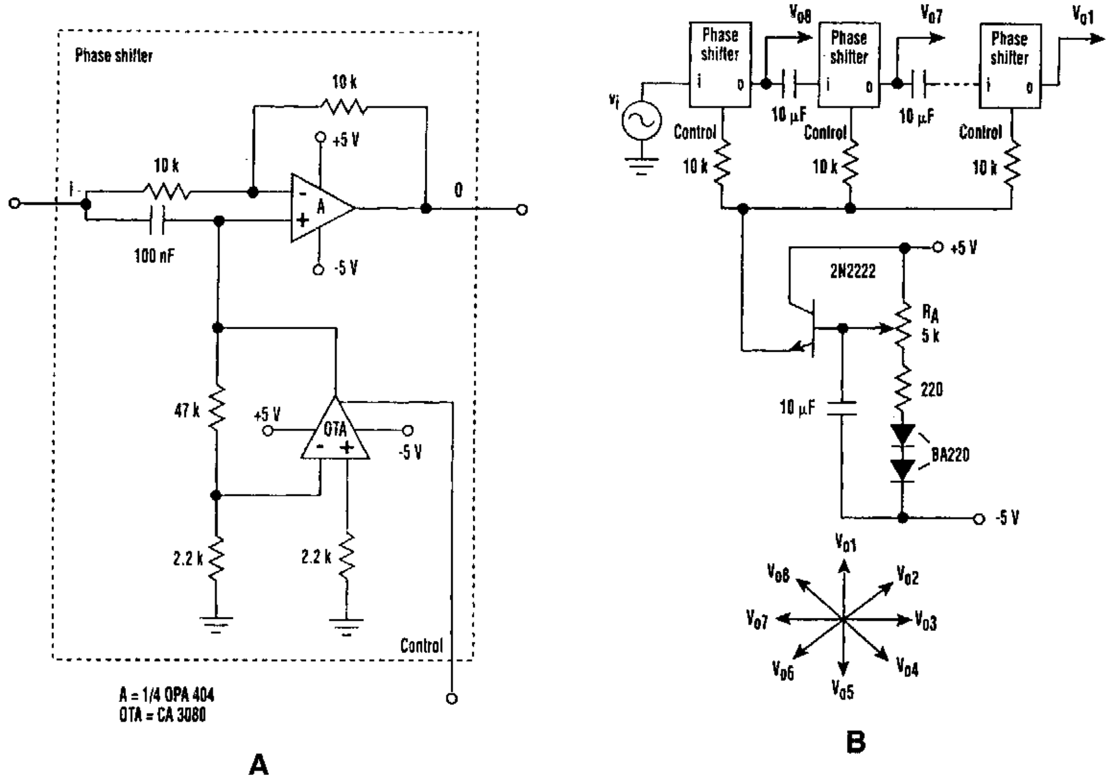
Q1: 2N2222, etc.

WILLIAM SHEETS

FIG. 74-2

This phase splitter uses a 2N2222 (or other general purpose npn transistor) to achieve outputs that are 180° out of phase.

PHASE SHIFTER WITH EIGHT OUTPUTS



ELECTRONIC DESIGN ***FIG. 74-3***

The circuit consists of eight cascaded identical cells, each cell being a dc-controlled active phase shifter. Because the dc control is common for all shifters, the circuit is adjusted by trimming R_A so that the phase difference between V_{01} and V_i is zero. As a result, each shifter will introduce a phase difference of exactly π/r. The eight signals for PSK are available at the op amps' outputs.

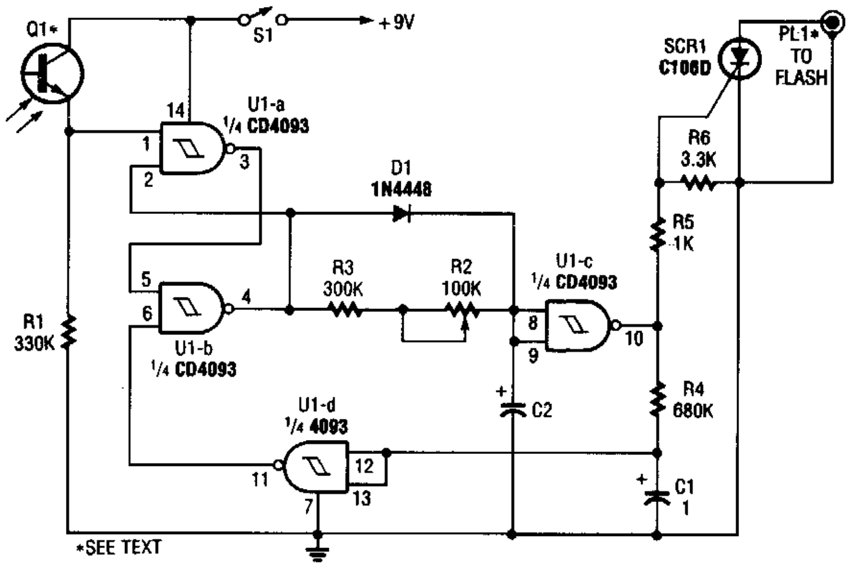
Phase accuracy is acceptable for 1%-tolerance resistors and 5%-tolerance 100-nF capacitors. Also, the amplitude of V_i (which is a 1700-Hz sine wave), should not exceed 1 V.

75

Photography Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

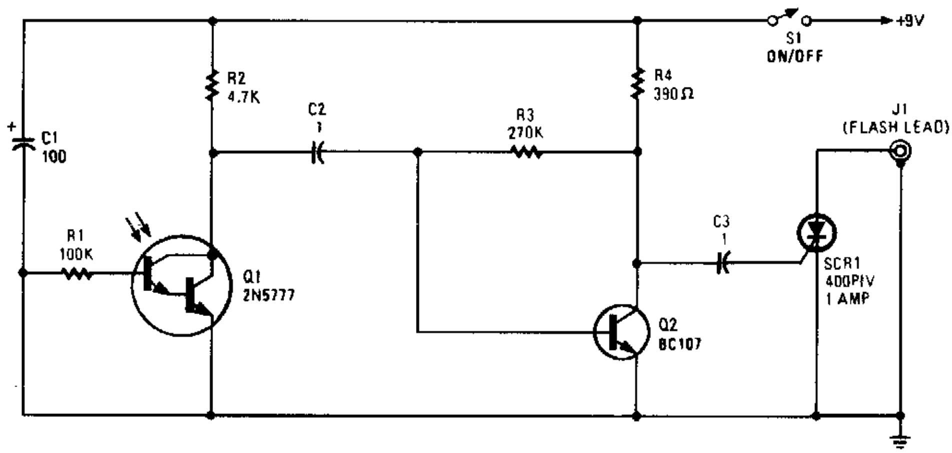
TIME-DELAY FLASH-TRIGGER CIRCUIT



1992 PE HOBBYIST HANDBOOK ***FIG. 75-1***

The circuit is built around a single 4093 quad 2-input NAND Schmitt trigger. Two gates from that quad package (U1-a and U1-b) are configured as a set-reset flip-flop.

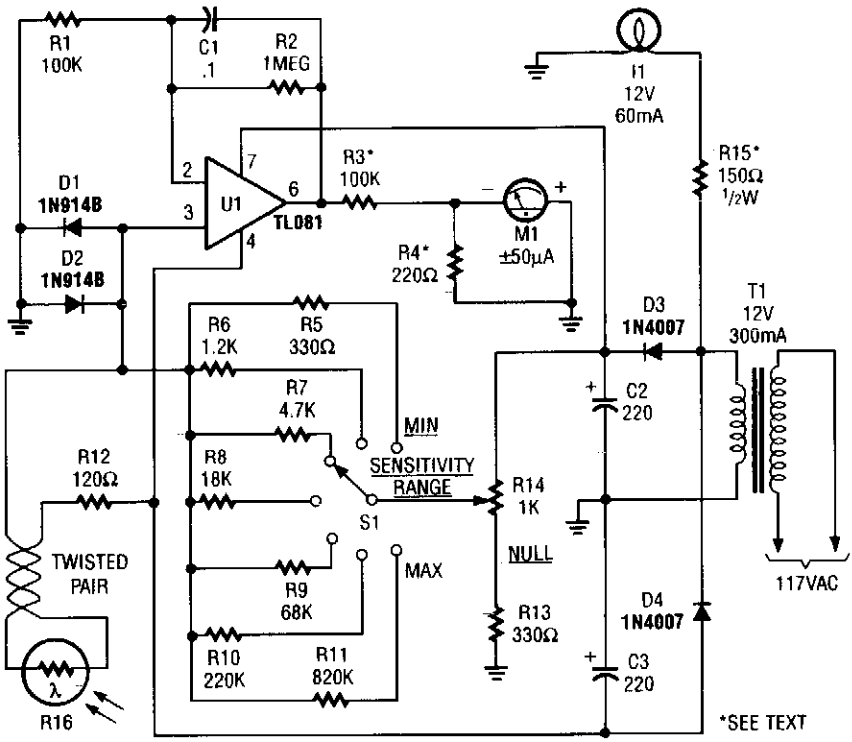
PHOTO FLASH SLAVE UNIT



POPULAR ELECTRONICS ***FIG. 75-2***

Phototransistor Q1 receives a light pulse from a photoflash unit. The pulse is ac-coupled to amplifier Q2. It then triggers SCR1, which triggers a flash unit that is connected to J1.

ENLARGING LIGHT METER



POPULAR ELECTRONICS

FIG. 75-3

Meter M1, a +/–50-μA zero-center D'Arsonval meter movement is driven by U1, a TL081 FET op amp, through R3. The gain of U1 is set at 11 by R1 and R2, while capacitor C1 is used to restrict the bandwidth of U1 to 1.6 Hz. Power for the circuit is derived from a simple dual-polarity 12-V power supply (consisting of T1, D3, D4, C2, and C3).

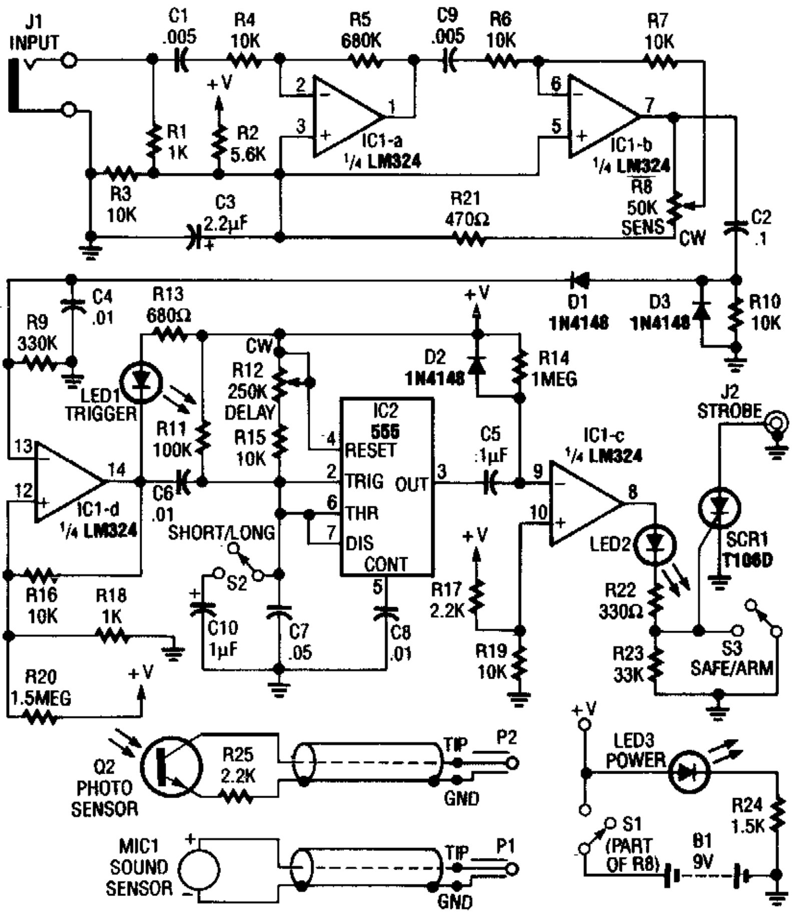
A light-dependent resistor (LDR), R16 (which is a semiconductor element whose resistance decreases as it is exposed to increasing illumination), is used as a light-sensing device. One end of R16 is connected to the negative supply rail through R12, and the other end is connected to pin 3 of U1, applying a negative current to U1. A variable (over a 4:1 range) positive current determined by the settings of R14 and S1 (and derived from the positive supply rail) is also fed to pin 3 of U1.

When the two currents (of opposite polarities) are equal, they cancel each other out, so effectively no current is applied to pin 3 of U1. With no current applied to pin 3, the output of U1 is zero and meter M1 registers accordingly, indicating a null. However, when light striking R16 causes its resistance to decrease, the current through the device increases, making the negative current greater than the positive current. Under that condition, the negative current causes the output of U1 to swing negative, causing the pointer to swing in the negative direction.

That indicates that the light intensity must be reduced by using a smaller lens opening on the enlarger (smaller f/stop). The opposite occurs if the light is too dim. Lamp 11, a 12-V 60-mA "grain of wheat" unit, is used to illuminate the meter scale, and R15 is used to limit the meter's illumination to a faint glow that is just bright enough so that the face of M1 can be plainly seen in a photo darkroom.

Resistors R3 and R4 should be selected for the meter used. With a dual supply of +/–12 V, U1 produces an output voltage of 10 V peak-to-peak. The resistance of R3 can be found by dividing the peak voltage (i.e., 10/2) by the full-scale meter current (in amps); i.e., R_3 = (10/2)/0.0005 = 100,000 Ω. R4, the shunt resistor, should be selected to have a value equal to the meter's internal resistance.

PHOTO STROBE

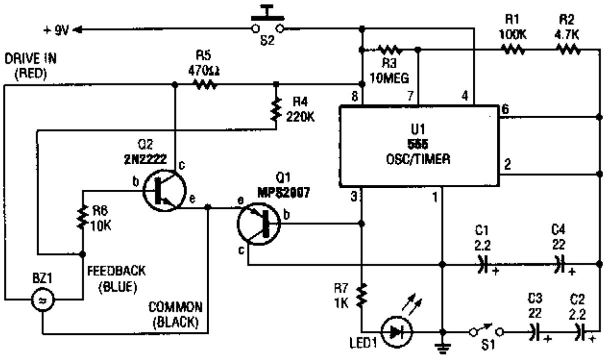


ELECTRONICS NOW

FIG. 75-4

Sound or light sensors connected to J2 produce a voltage that is amplified by IC1-a and IC1-b. A positive trigger voltage that is developed by D1 and D3 and amplified by IC1-d, drives IC2 and IC1 to trigger SCR1. SCR1 is connected to a strobe. This device is handy for photographic purposes to take pictures of events that involve sound, such as impacts, etc.

DARKROOM TIMER



1991 PE HOBBYIST HANDBOOK

FIG. 75-5

The electronic darkroom timer is built around a 555 oscillator/timer, a pair of general-purpose transistors, a buzzer, and an LED. The 555 (U1) is configured as an astable multivibrator (free-running oscillator). The frequency of the oscillator is determined by the values R_1 through R_3 and C_1 through C_4.

Switch S1 is used to divide the capacitor network to vary the time interval between beeps; when S1 is closed, the circuit beeps at intervals of 30 seconds. With S1 closed, it beeps at 15-second intervals.

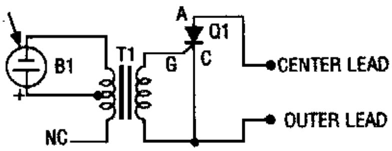
When power is applied to the circuit (by closing switch S2), the output of U1 at pin 3 is initially high. That high is applied to the base of transistor Q1 (an MPS2907 general-purpose pnp device), keeping it turned off. That high is also applied to the anode of LED1 (which is used as a power on indicator) through resistor R7, turning it on.

Timing capacitors C1 through C5 begin to charge through timing resistors R1 through R3. dc voltage is applied to BZ1's driver input through R5 and to its feedback terminal (through R4), which is also connected to Q2's base terminal. The $V+$ voltage that applied to Q2's base causes it to turn on, tying BZ1's common terminal high.

When the timing capacitors are sufficiently charged, a trigger pulse is applied to pin 2 (the trigger input) of U2, causing U1's output to momentarily go low. This causes LED1 to go out and transistor Q1 to turn on. That, in turn, grounds the common lead of buzzer BZ1, causing BZ1 to sound. Afterward, the output of U1 returns to the high state, turning off Q1, and turning on LED1, until another time interval has elapsed and the process is repeated.

The circuit is powered by a 9-Vac adapter, which plugs into a standard 117-V household outlet. Because the circuit draws only about 10 to 15 mA, a 9-V alkaline transistor-radio-battery can also be used to power the circuit.

PHOTO STROBE SLAVE TRIGGER

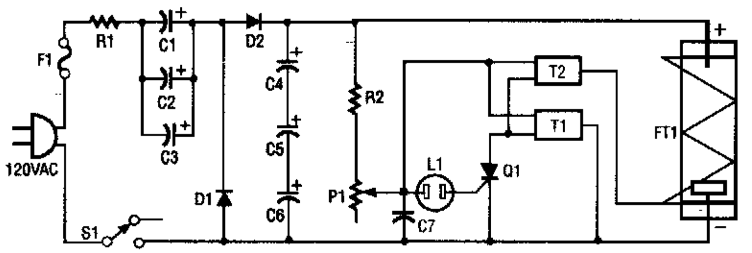


1991 PE HOBBYIST HANDBOOK

FIG. 75-6

The photo strobe slave trigger circuit uses a solar cell and an SCR to flash any strobe when you trigger your "master" strobe. The tiny solar cell produces a very small voltage when light falls on its surface.

STROBE LIGHT



C1,C2,C3	10 µF 160V Electrolytic Capacitor
C4,C5,C6	160 µF 200V Electrolytic Capacitor
C7	0.5 µF 250V Mylar Capacitor
D1, D2	1N4004 Diodes
F1	1 Amp Pigtail Fuse
FT1	Giant Xenon Strobe Tube
L1	Neon Lamp
P1	10 Meg Potentiometer
Q1	106D1 SCR
R1	20 ohm 10 Watt Power Resistor
R2	270K 1/4 Watt Resistor
S1	Slide Switch
T1, T2	Trigger Coil

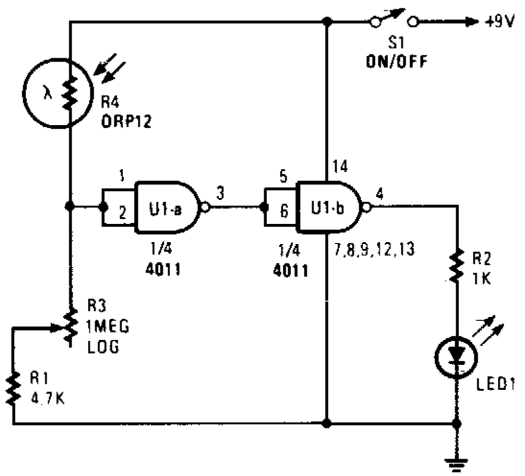
1991 PE HOBBYIST HANDBOOK ***FIG. 75-7***

This strobe light operates from standard 120-Vac power. R1 limits the amount of current applied to the voltage doubler stage, which is comprised of C1, C2, C3, D1, D2, C4, C5, and C6. Capacitors C1, C2, and C3 are connected in parallel and form a capacitance of 30 µF at 160 V. Capacitors C4, C5, and C6 are connected in series and form an equivalent capacitor of about 53 µF at 480 V. Diodes D1 and D2 not only rectify the ac voltage, but also complete the voltage doubler stage, which converts the incoming 120 Vac to the appropriately 300 V that are required by the xenon strobe tube.

The next stage of the circuit is the neon relaxation oscillator and trigger stage. This stage is made up of R2, P1, C7, L1, Q1, T1, and T2. As the storage capacitor (made up of C4, C5, and C6) reaches its full-capacity charge, the voltage divider (made up of R2 and P1) applies voltage to capacitor C7. As C7 charges up, it reaches a threshold voltage level, SCR Q1. When Q1 has a positive pulse on its gate, it fires (causes a short from anode to cathode). That firing action discharges most of the energy stored in C7 into trigger transformers T1 and T2 (which have secondaries connected in series to developer 8 kV). The frequency of the 8-kV pulses is determined by the setting of P1 and the value of C_7. Because C7 is a fixed capacitor, only the setting of P1 adjusts the flash rate in this circuit.

As soon as an 8-kV pulse is applied from the secondary of T2 (trigger wire) to the trigger lead of FT1, it discharges storage capacitors C4, C5, and C6, which causes it to ionize (flash). The cycle then repeats itself until the power is removed from the circuit board by turning "off" S1 or removing the line cord.

ENLARGER EXPOSURE METER



POPULAR ELECTRONICS

FIG. 75-8

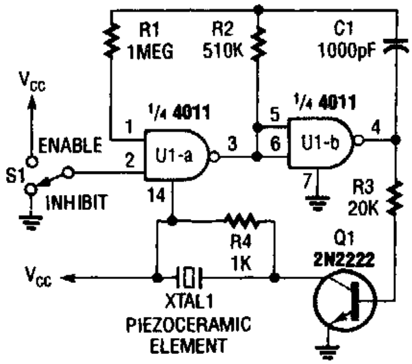
Two gates of a 4011 are used as a comparator. When the resistance of R4 decreases the voltage at pin 1 and 2 increases, producing a logic zero at pin 3, causing pin 4 to go high and activating the LED. R3 is calibrated in light units, or seconds exposure time. To calibrate, set pot R3 so as to just be on the LED ON/OFF threshold. With a light level that is suitable to correctly expose a photographic print, use a known enlarger and a known negative.

76

Piezo Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

CMOS PIEZO DRIVER



POPULAR ELECTRONICS

FIG. 76-1

A CMOS-gate and transistor buffer can be used as an effective driver for a piezoelectric transducer.

CMOS PIEZO DRIVER USING 4049

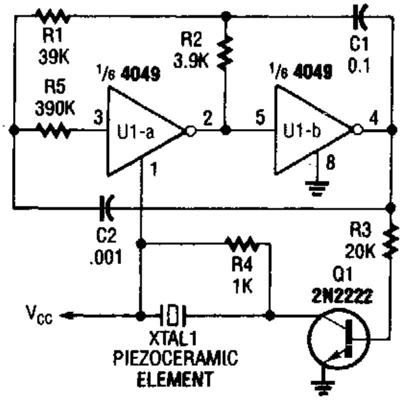
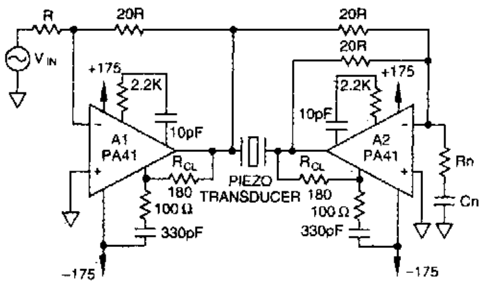


FIG 76-2

This circuit uses a 4049 IC to drive a 2N2222 switching transistor. The transistor drives crystal 1 a piezo transducer.

PIEZO DRIVER

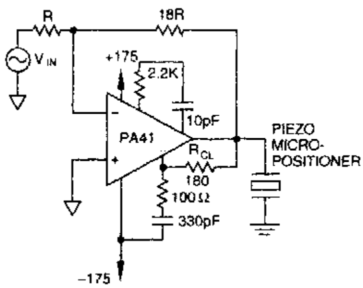


ELECTRONIC DESIGN

FIG. 76-3

Using a PA41 from Apex Microtechnology, this monolithic amplifier is capable of 350-V operation and delivers 660 V p-p in a bridge circuit.

PIEZO MICROPOSITIONER DRIVER

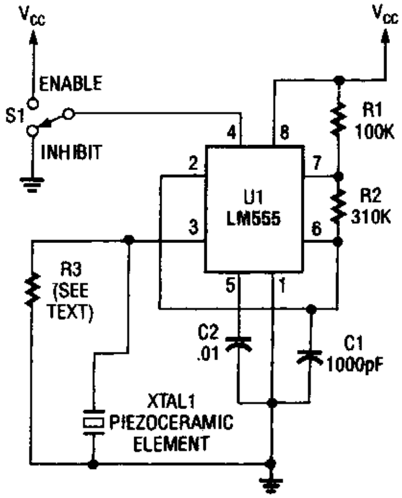


ELECTRONIC DESIGN

FIG. 76-4

The PA41 from Apex Microtechnology is used here to drive a piezoelectric micropositioner. The drive voltage is less than 20 V p-p at input.

555 OSCILLATOR FOR DRIVING A PIEZO TRANSDUCER



POPULAR ELECTRONICS **FIG. 76-5**

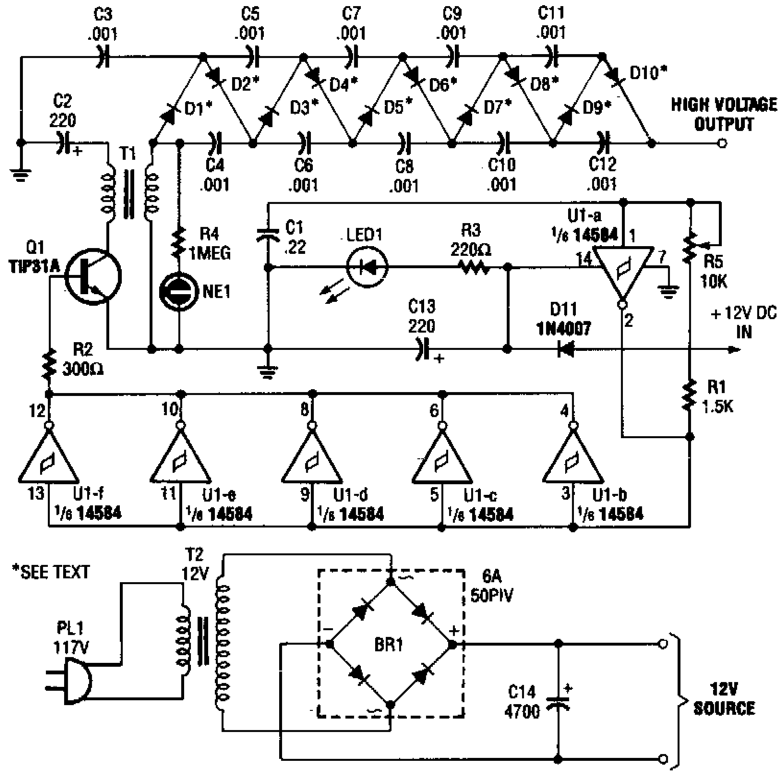
A 555-timer oscillator is perhaps one of the most popular circuits for driving a piezoelectric transducer.

77

Power Supply Circuits—High Voltage

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

HIGH-VOLTAGE dc GENERATOR



1990 PE HOBBYIST HANDBOOK

FIG. 77-1

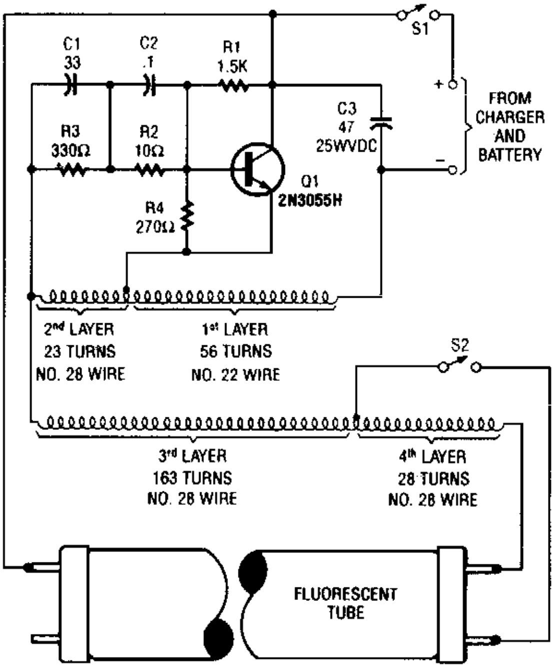
In the miniature high-voltage dc generator, the input to the circuit, taken from a 12-Vdc power supply, is magnified to provide a 10,000-Vdc output causing a pulsating signal, of opposite polarity, to be induced in T1's secondary winding.

The pulsating dc output at the secondary winding of T1 (ranging from 800 to 1000 V) is applied to a 10-stage voltage-multiplier circuit, which consists of D1 through D10, and C3 through C12. The multiplier circuit increased the voltage 10 times, producing an output of up to 10,000 Vdc. The multiplier accomplishes its task by charging the capacitors (C3 through C12); the output is a series addition of the voltages on all the capacitors in the multiplier.

In order for the circuit to operate efficiently, the frequency of the square wave, and therefore the signal applied to the multiplier, must be considered. The output frequency of the oscillator (U1-a) is set by the combined values of R_1, R_5, and C_1 (which with the values specified is approximately 15 kHz). Potentiometer R5 is used to fine tune the output frequency of the oscillator. The higher the frequency of the oscillator, the lower the capacitive reactance in the multiplier.

Light-emitting diode LED1 serves as an input-power indicator, and neon lamp NE1 indicates an output at the secondary of T1. A good way to get the maximum output at the multiplier is to connect an oscilloscope to the high-voltage output of the multiplier, via a high-voltage probe, and adjust potentiometer R5 for the maximum voltage output.

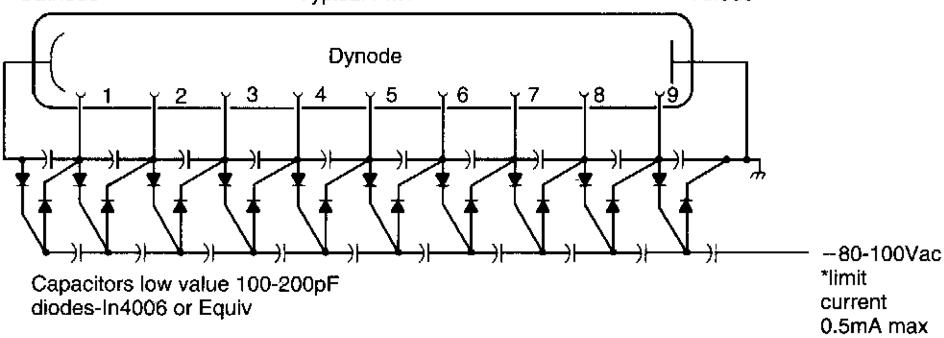
FLUORESCENT TUBE POWER SUPPLY



POPULAR ELECTRONICS ***FIG. 77-2***

A 2N3055 oscillator (Q1) drives a homemade transformer, wound on a 5/16 × 1⅞" ferrite rod. S2 is used as a filament switch and it can be eliminated, if desired. A 20-W fluorescent tube is recommended. The supply is 12 V.

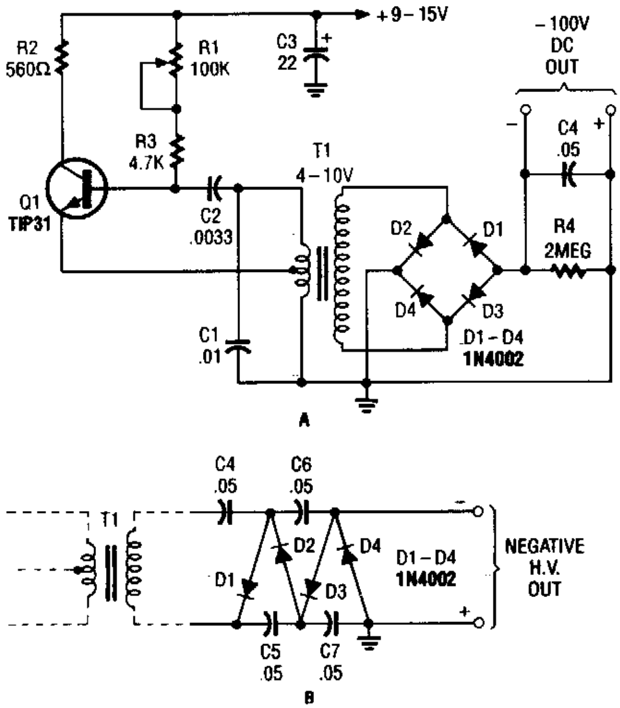
PHOTOMULTIPLIER SUPPLY



73 AMATEUR RADIO TODAY ***FIG. 77-3***

A Cockcroft-Walton voltage multiplier supplies the stepped voltage required for the dynodes of the PMT without the power-wasting voltage-divider resistor string that is traditionally used.

NEGATIVE VOLTAGE SUPPLY

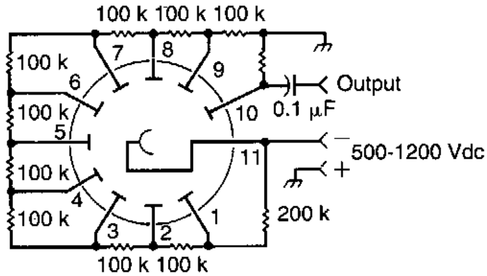


POPULAR ELECTRONICS

FIG. 77-4

The combination Hartley oscillator/step-up transformer shown in A can generate significant negative high voltage, especially if the voltage output of the transformer is multiplied by the circuit.

PHOTOMULTIPLIER CIRCUIT

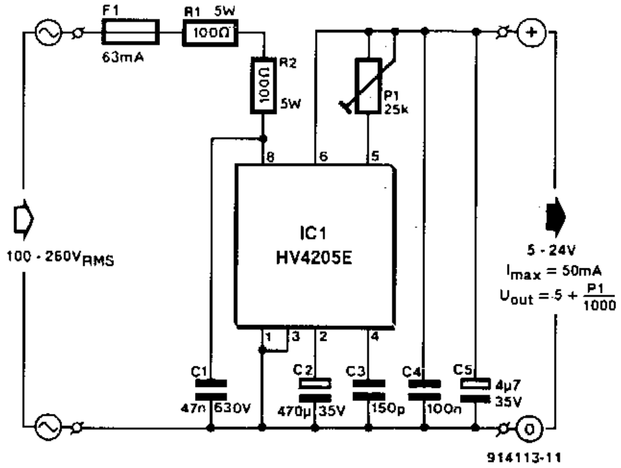


73 AMATEUR RADIO TODAY

FIG. 77-5

This circuit is typical of the way that a photomultiplier tube is used. The circuit shown is ac coupled, but if dc coupling is needed, the capacitor can be omitted and a suitable interfacing method used. A typical tube is the widely available 931/931A.

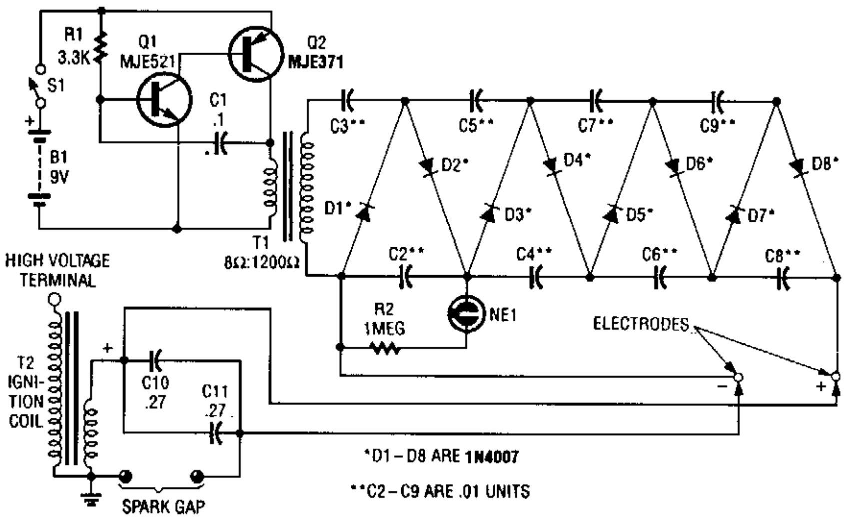
SINGLE-CHIP dc SUPPLY FOR 120–TO 240-Vac OPERATION



ELEKTOR ELECTRONICS **FIG. 77-6**

Direct derivation of 5 to 24 Vdc from ac mains, without a transformer is possible with this circuit. Note that a direct mains connection to the dc output exists. *Suitable safety precautions must be taken.*

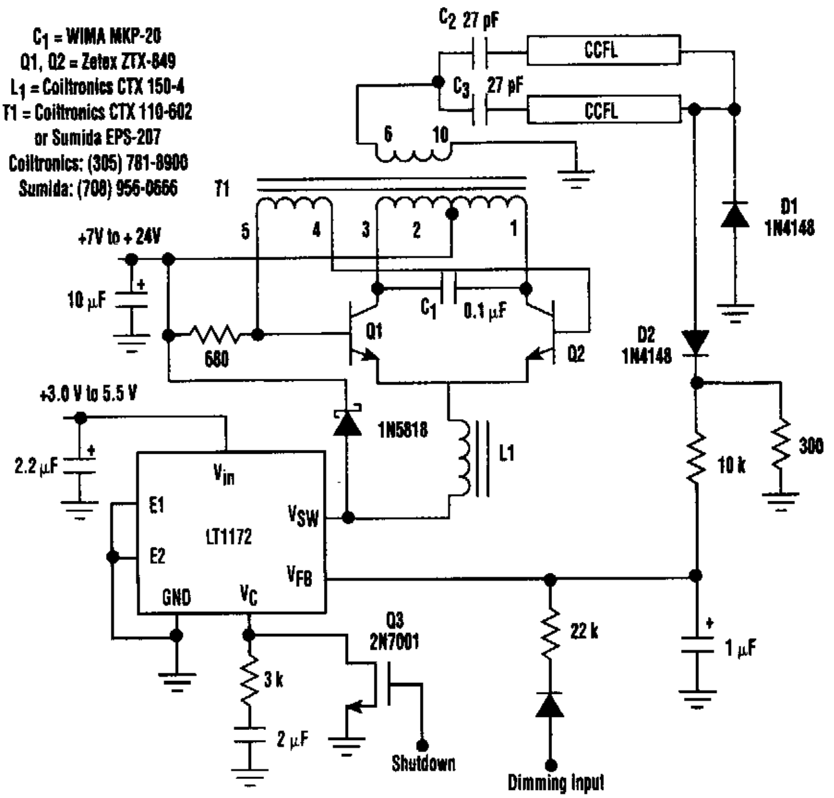
HIGH-VOLTAGE SUPPLY



POPULAR ELECTRONICS **FIG. 77-7**

This circuit uses a transistor oscillator and a voltage multiplier to charge C10 and C11 to a high voltage. When the spark gap breaks down, T2 produces a high-voltage pulse via the capacitance discharge of C10 and C11 into its primary. T2 is an auto ignition coil.

COLD-CATHODE FLUORESCENT-LAMP POWER SUPPLY



ELECTRONIC DESIGN *FIG. 77-8*

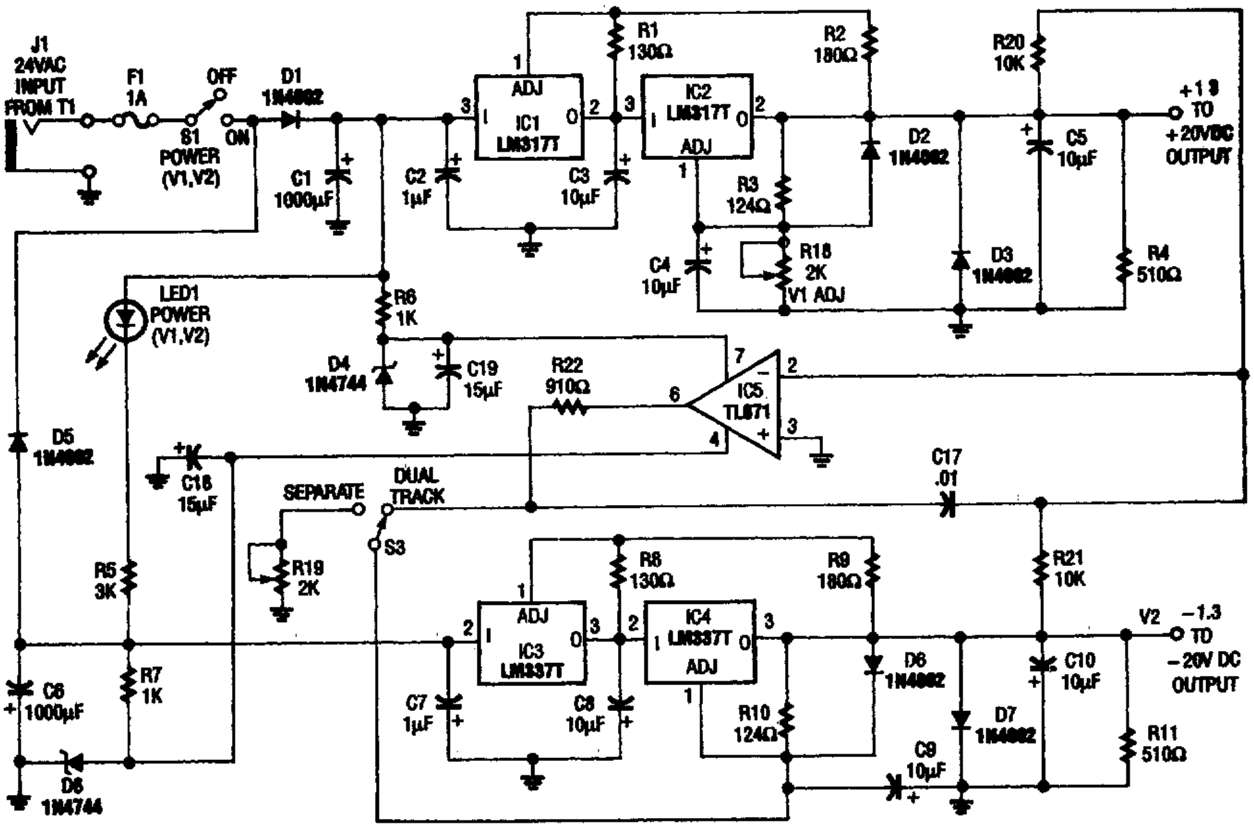
This circuit is a 92%-efficient power supply for cold-cathode fluorescent lamps (CCFLs), which are used to backlight LCD in portable equipment. The efficiency depends heavily on the component types, particularly C1, Q1, Q2, L1, and T1, whose manufacturers are noted.

78

Power Supply Circuits—Low Voltage

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

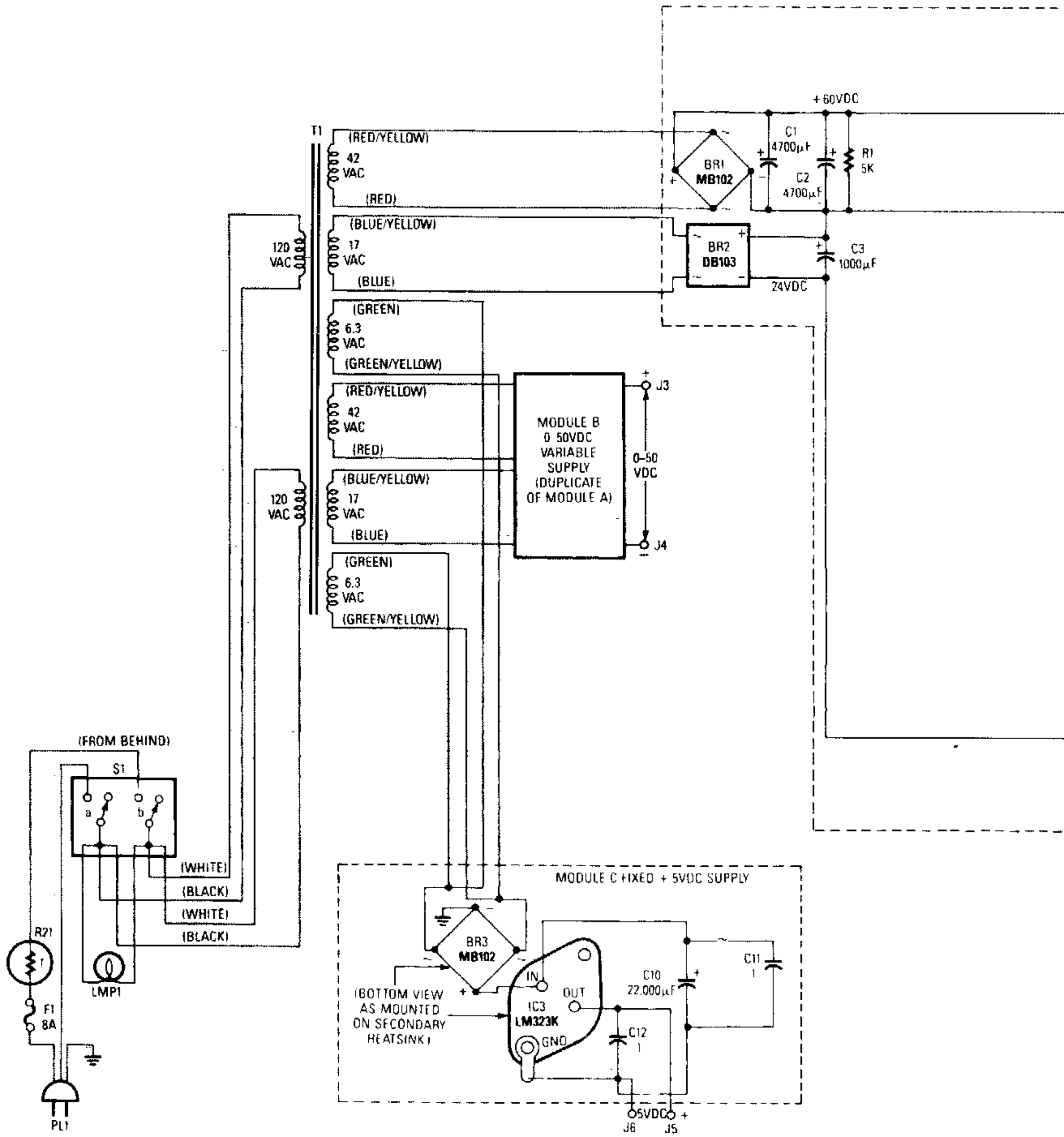
TRACKING DOUBLE-OUTPUT BIPOLAR SUPPLY



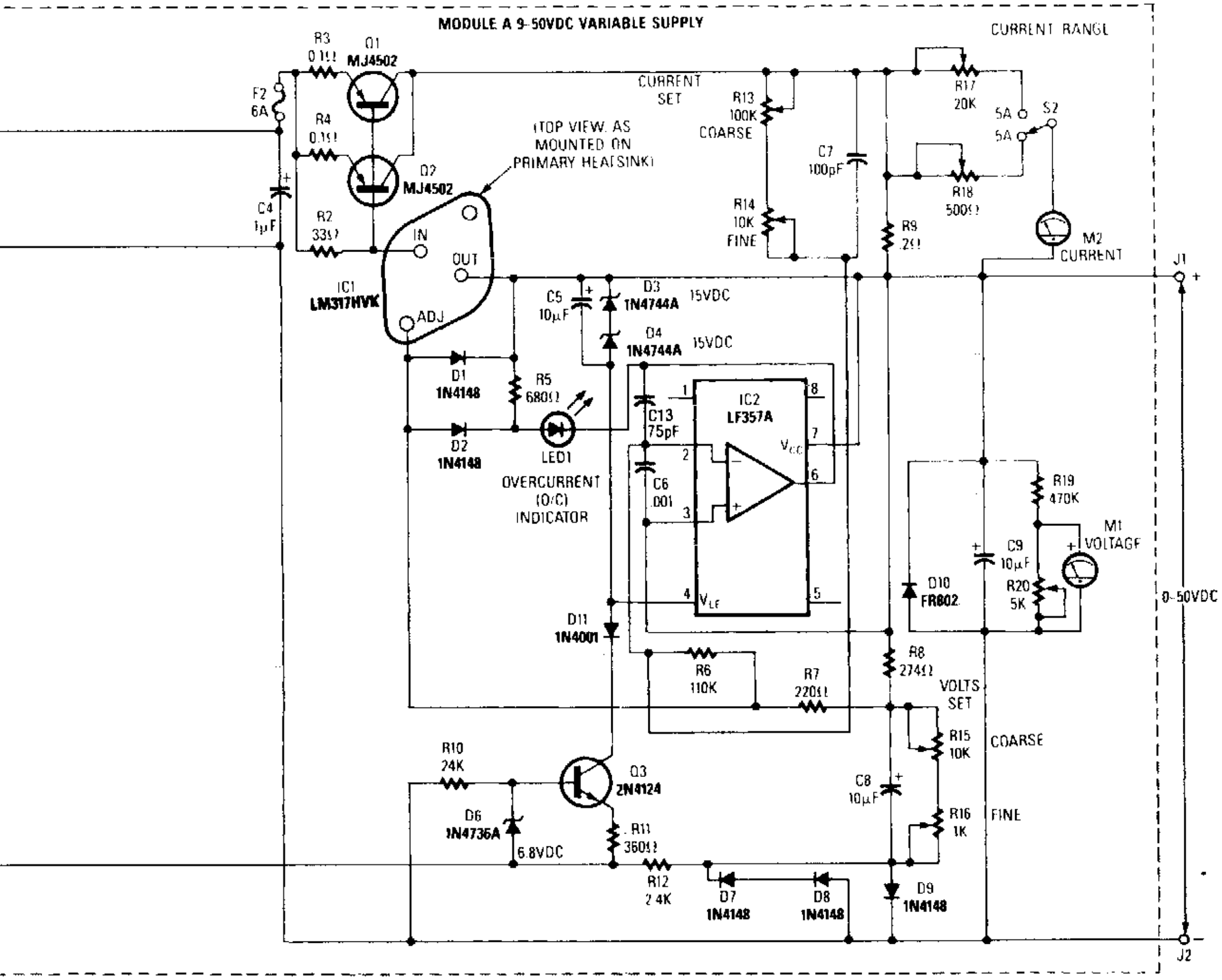
ELECTRONICS NOW

FIG. 78-1

This circuit is useful for a bench supply in the lab. Separate or tracking operation is possible. The regulators should be properly heatsinked. T1 is a 24-Vac wall transformer of suitable current capacity.



+60VDC
T1
(RED/YELLOW)
42 VAC
(RED)
BR1 MB102
C1 4700µF
C2 4700µF
R1 5K
(BLUE/YELLOW)
17 VAC
(BLUE)
BR2 DB103
C3 1000µF
24VDC
120 VAC
(GREEN)
6.3 VAC
(GREEN/YELLOW)
MODULE B 0 50VDC VARIABLE SUPPLY (DUPLICATE OF MODULE A)
J3
0–50 VDC
J4
(FROM BEHIND)
S1
(WHITE)
(BLACK)
R21
LMP1
F1 8A
PL1
MODULE C FIXED + 5VDC SUPPLY
BR3 MB102
(BOTTOM VIEW AS MOUNTED ON SECONDARY HEATSINK)
IN
OUT
GND
IC3 LM323K
C10 22,000µF
C11 1
C12 1
5VDC
J6
J5



The value of the design lies in the use of IC1, an LM317HVK adjustable series-pass voltage regulator, for broad-range performance remainder supplies voltage-setting and current-limiting functions. The input to IC1 comes from the output of BR1, which is filtered by C1 and C2 to about +60 Vdc, and the input for current-sense comparator IC2 comes from BR2, which also acts as a negative bias supply for regulation down to ground. The output voltage is determined by:

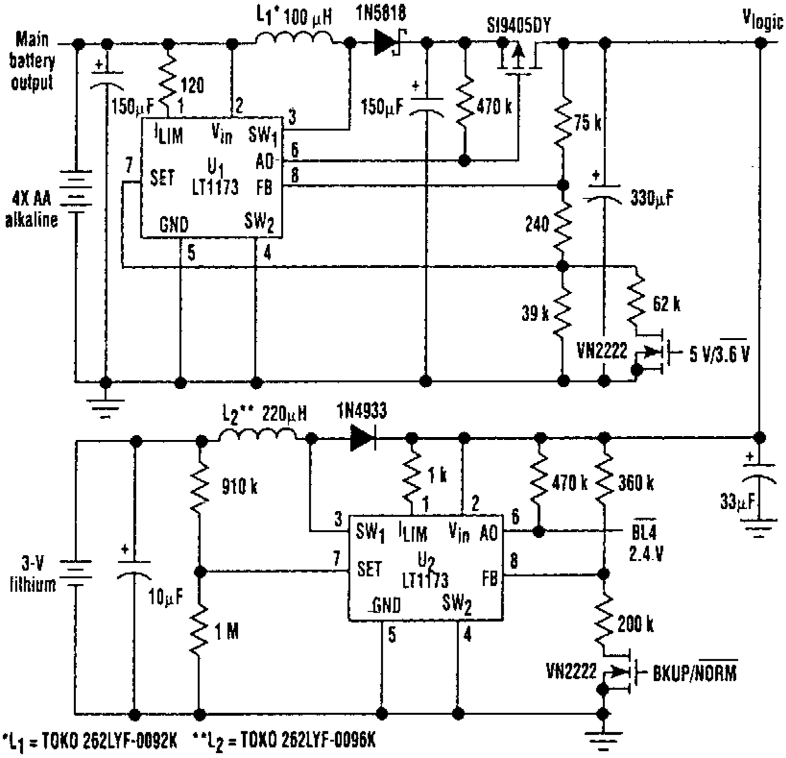
$$(V_{OUT} - 1.25 + 1.3)/(R_{15} + R_{16}) = 1.25/R_8.$$

Thus, the maximum value from each variable supply board is:

$$V_{OUT} = (1.25/R_8) \times (R_{15} + R_{16}) = 50.18 \text{ Vdc}.$$

FIG. 78-2

+5 V/+3.6 V FROM 4 AA CELLS



ELECTRONIC DESIGN

FIG. 78-3

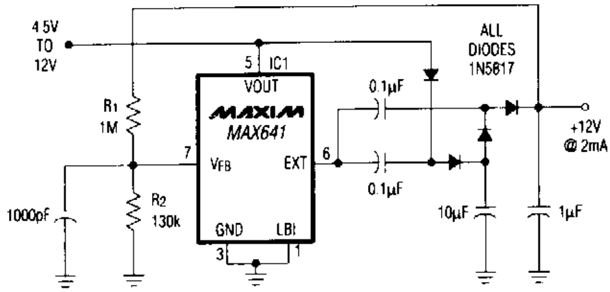
With this unique logic-power-converter design (see the figure), a switchable 3.6 or 5 V at 200 mA can be attained by using four AA cells. The supply incorporates a MOSFET switch that can switch to a lithium backup battery, providing a 3.4-V output when the main battery is dead or removed. The supply consumes only 380 µA under no-load conditions.

The circuit operates in a somewhat novel mode as a step-up/step-down converter. When the cells are fresh (from about 6 V to about 5.2 V), the LT1173's gain block drives the p-channel MOSFET, which turns the circuit into a linear voltage regulator. This might seem inefficient, but the batteries are quick to drop from 6 V to 5 V. With a 5-V input, the efficiency (for the 3.6-V output) is 3.6/5 or 72%, which is reasonable. As the battery-pack drops in voltage, efficiency increases, reaching greater than 90% with a 4.2-V input.

At a point below a 4-V input, the circuit switches to step-up mode. This mode squeezes the batteries for all of their available energy. In this case, efficiency runs between 83% at approximately a 4-V input to 73% at a 2.5-V input.

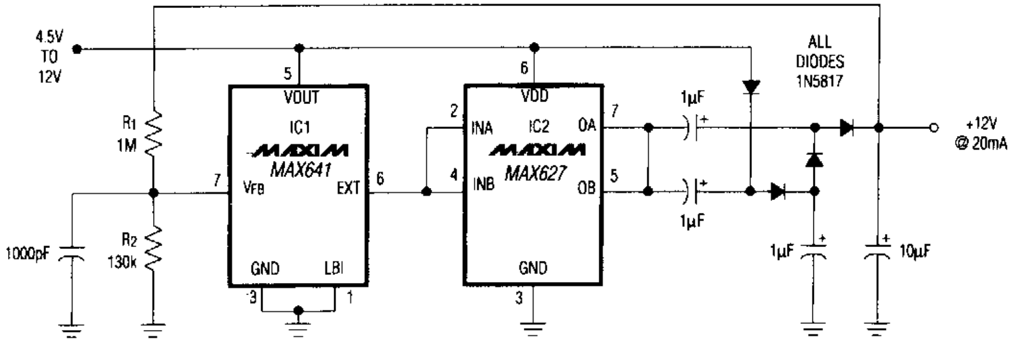
The supply can deliver 200 mA over its entire operational range. In its linear mode of operation, the supply has no current spikes that, because of the fairly high internal resistance of the alkaline cells, can reduce battery life. The topology delivers over 9.3 hours of 3.6-V 200-mA output power, compared to just 7 hours using the traditional flyback topology that is used in other designs.

INDUCTORLESS SWITCHING REGULATOR



Substituting the diode-capacitor network shown for an inductor allows this switching-regulator IC to deliver 2mA at comparable line and load regulation, with somewhat reduced efficiency.

A



Introducing an MOS driver (IC2) enables the Figure 1 circuit to deliver as much as 20mA.

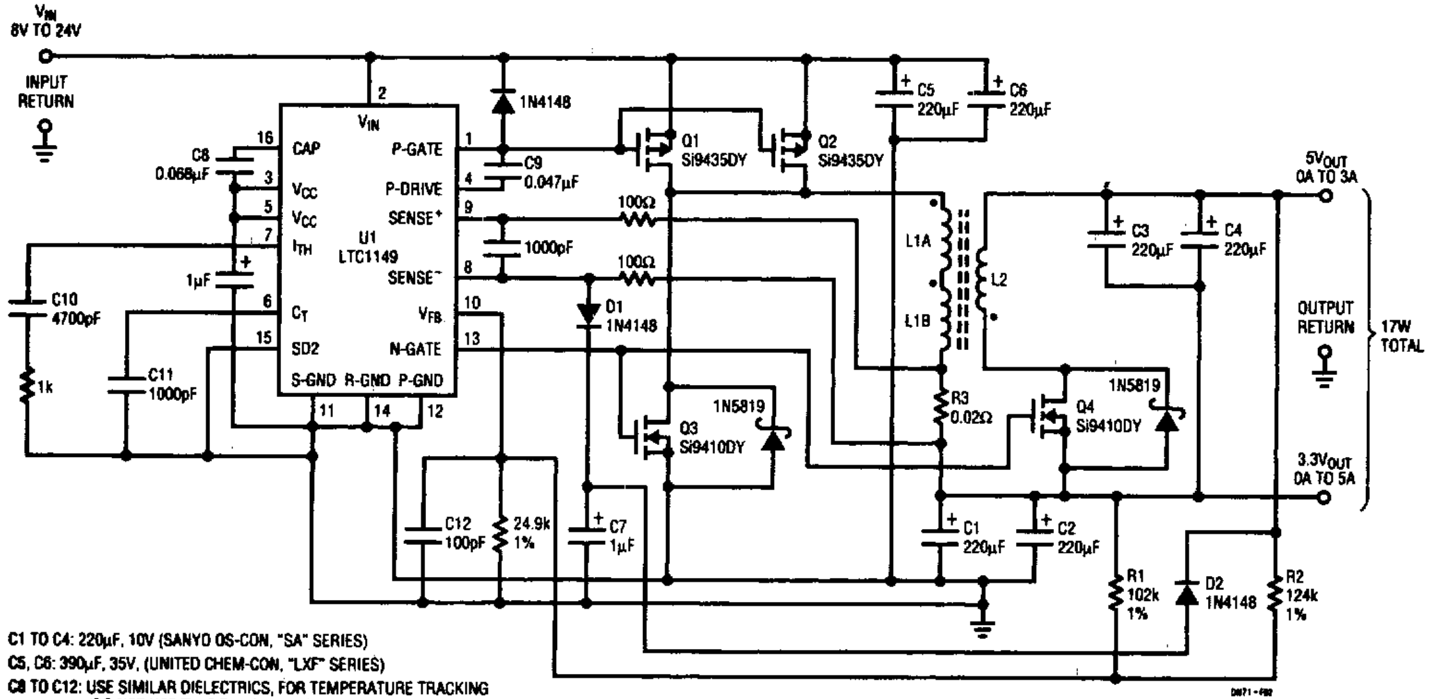
B

MAXIM ENGINEERING JOURNAL ***FIG. 78-4***

In conventional applications, switching-regulator ICs regulate V_{OUT} by controlling the current through an external inductor. The IC in A, however, driving a diode-capacitor network in place of the inductor, offers comparable performance for small loads. The network can double, triple, or quadruple the input voltage.

Feedback from the R1/R2 voltage divider enables IC1 to set the regulated-output level. (As shown, the circuit derives 12 V from a 5- to 12-V input and provides as much as 2 mA of output current.) Adding a noninverting MOS driver (B) boosts the available output current to 20 mA. Substituting the diode-capacitor network shown for an inductor allows this switching-regulator IC to deliver 2 mA at comparable line and load regulation, with somewhat reduced efficiency.

SINGLE LTC POWER SUPPLY



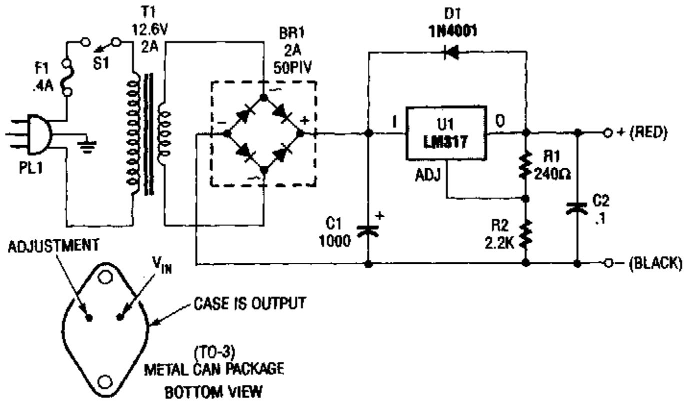
C1 TO C4: 220µF, 10V (SANYO OS-CON, "SA" SERIES)
C5, C6: 390µF, 35V, (UNITED CHEM-CON, "LXF" SERIES)
C8 TO C12: USE SIMILAR DIELECTRICS, FOR TEMPERATURE TRACKING
(DO NOT USE TYPE Z5U CERAMIC CAPACITORS)
R3: 0.02Ω, NONINDUCTIVE. IRC TYPE MSM-1-0.02Ω-±5%.
L1A, L1B, L2: SEE TEXT
UNMARKED RESISTORS ARE 0.25W, 5%

LINEAR TECHNOLOGY

FIG. 78-5

One LTC 1149 synchronous switching regulator can deliver both 3.3- and 5-V outputs. The design's simplicity, low cost, and high efficiency make it a strong contender for portable, battery-powered applications. The circuit described accepts input voltages from 8 to 24 V, to power any combination of 3.3-V and 5-V loads totalling 17 W or less. For input voltages in the 8-V to 16-V range, the LTC1148 may be used, reducing both quiescent current and cost.

CONFIGURABLE POWER SUPPLY

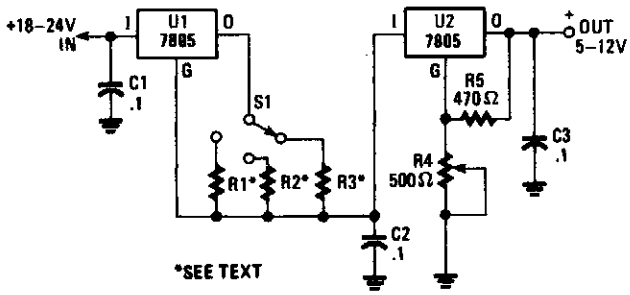


POPULAR ELECTRONICS ***FIG. 78-6***

The adjustable supply can easily be reconfigured by altering the value of V_2 and beefing up some other components, as is necessary.

The output voltage is given by $V_{OUT} = 1.25\ (1 + R_2/R_1)$. R_2 can be changed, as is necessary.

COMBINATION VOLTAGE AND CURRENT REGULATOR



POPULAR ELECTRONICS ***FIG. 78-7***

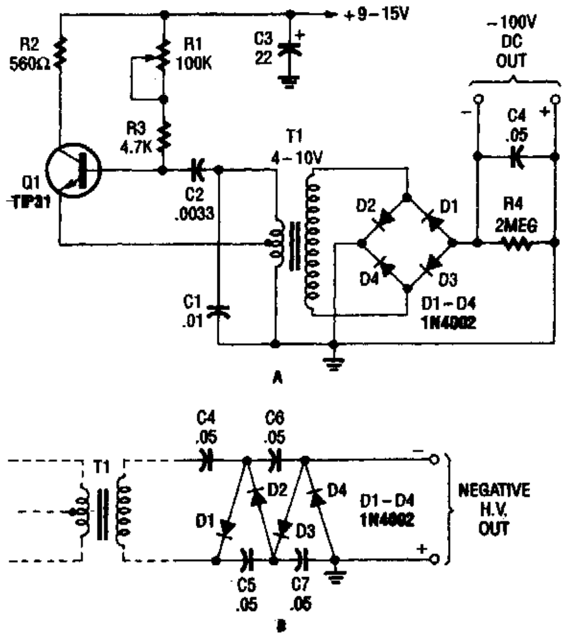
This voltage-regulator/current-limiter combination can be made from two 7805 regulators as shown. R1, R2, and R3 should be selected for a 5-V drop at the maximum allowable current limit. S1 selects one of the three current values. Do not forget that U1 requires 5 mA to operate and this means that the minimum current limit setting should be 10 mA or more ($R_1 = 1.25\ k\Omega$). Resistor values are as follows:

$$R_x\ (k\Omega) = \frac{5\ \text{volts}}{(\textit{current limit}\ \text{mA} - 5\ \text{mA})}$$

For 100 mA,

$$R_x = \frac{5}{100-5} = \frac{5}{95}\ k\Omega \text{ or } 52.5\ \Omega$$

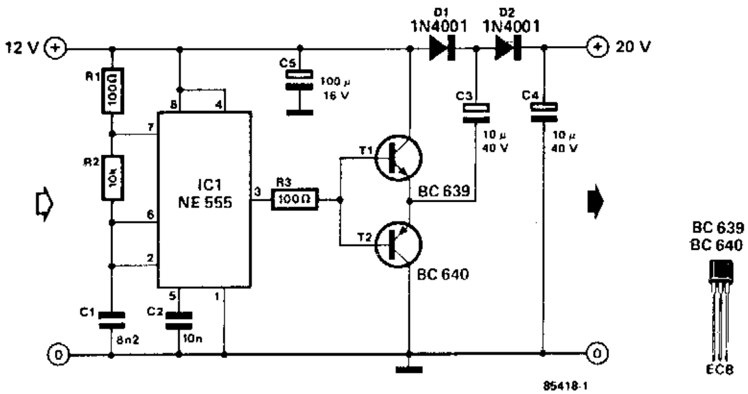
HV POWER SUPPLY WITH 9-TO 15-Vdc INPUT



POPULAR ELECTRONICS *FIG. 78-8*

The combination Hartley oscillator/step-up transformer shown in A can generate significant negative high voltage, especially if the voltage output of the transformer is multiplied by the circuit in B.

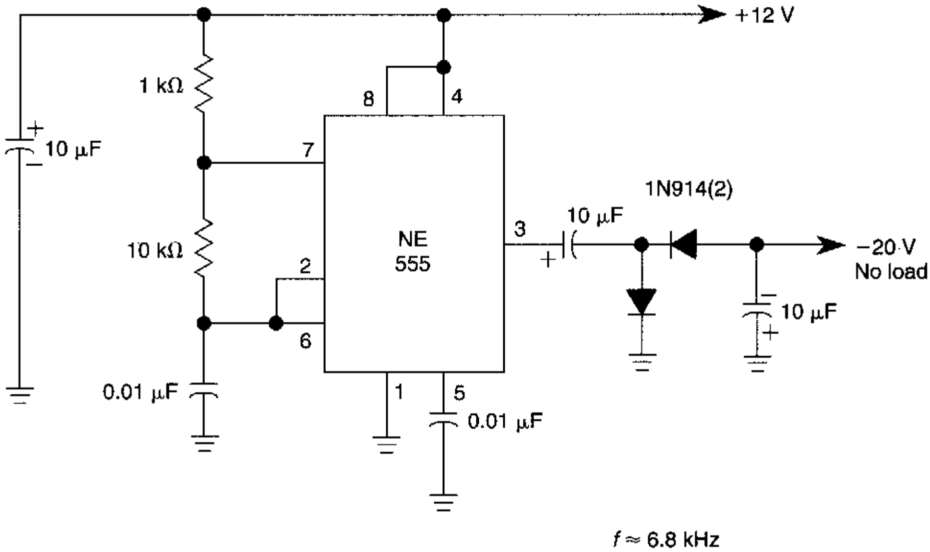
INDUCTORLESS POWER SUPPLY CONVERTER



303 CIRCUITS *FIG. 78-9*

Using a 555 timer and voltage doubler, this circuit will supply ≥50mA at 20 Vdc. T1 and T2 act as power amplifiers to drive the voltage doubler. Frequency of operation is approximately 8.5 kHz.

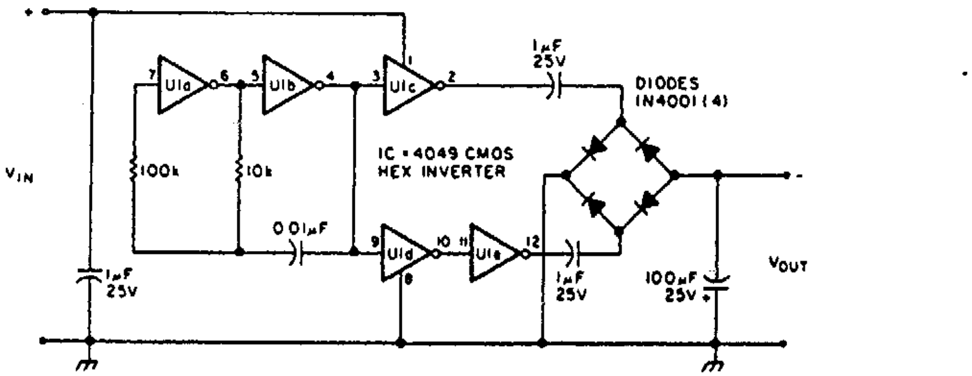
SIMPLE NEGATIVE SUPPLY FOR LOW-CURRENT APPLICATIONS



WILLIAM SHEETS ***FIG. 78-10***

This dc negative-voltage generator based on the 555 produces a negative output voltage equal to approximately 2x the dc supply voltage.

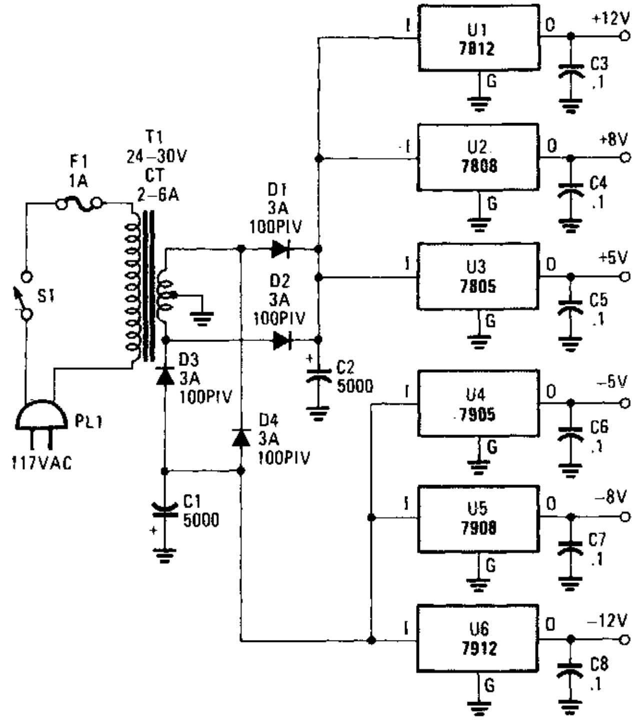
INVERTING POWER SUPPLY



73 AMATEUR RADIO TODAY ***FIG. 78-11***

This circuit will provide a negative dc voltage that is approximately equal to the positive input voltage at no load and about 3 V less at 10 mA load. V_{IN} is from +5 to +15 Vdc. Do not exceed 15 V or U1 might be damaged.

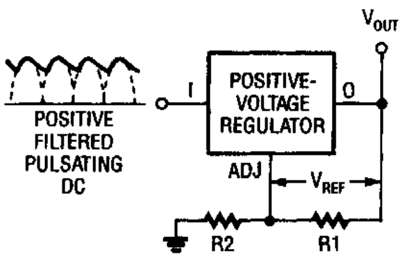
MULTIVOLTAGE POWER SUPPLY



POPULAR ELECTRONICS *FIG. 78-12*

This dual-polarity, multivoltage power supply can be built for a very small investment. The circuit is built around 78XX and 79XX series 1-A voltage regulators, four 3-A diodes, a 24–30-V 2–6-A transformer, and eight filter capacitors.

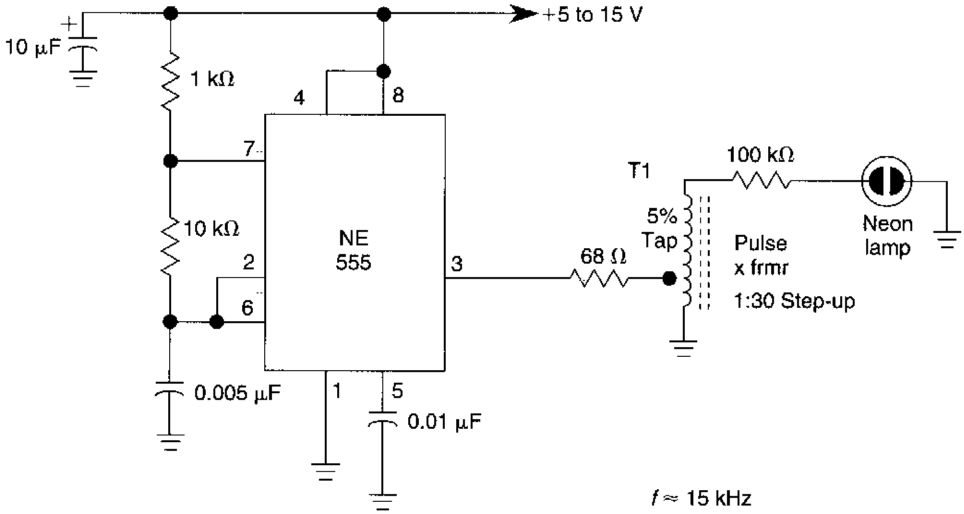
CURRENT-LIMITING REGULATOR



Floating adjustable regulators can be used as current limiters. Resistor R1 programs the current flowing through R2.

1993 ELECTRONICS HOBBYISTS HANDBOOK *FIG. 78-13*

NEON LAMP DRIVER FOR 5- TO 15-V SUPPLIES

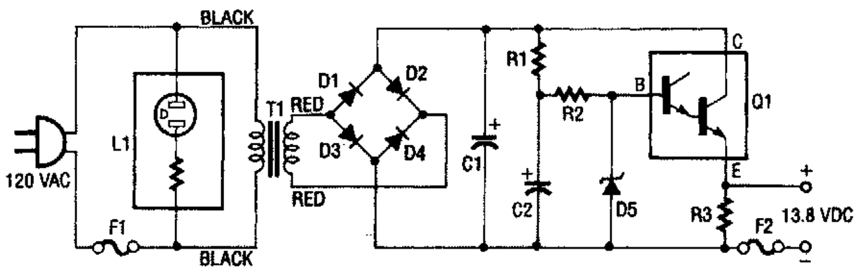


WILLIAM SHEETS

FIG. 78-14

This neon-lamp driver based on the 555 T1 can be wound on an old TV flyback transformer core.

13.8-Vdc 2-A REGULATED POWER SUPPLY

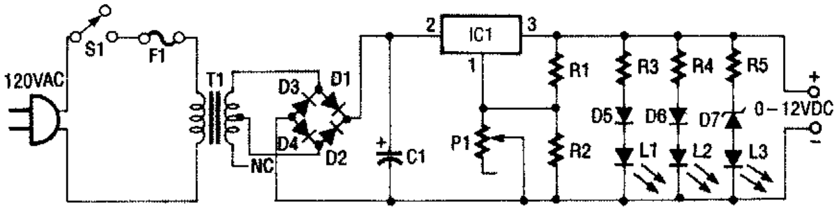


1991 PE HOBBYIST HANDBOOK

FIG. 78-15

This regulated power supply consists of step-down transformer T1, a full-wave rectifier bridge (D1 through D4), and a filtering regulator circuit made up of C1, C2, R1, R2, R3, D5, and Q1. When 120 Vac is provided, the neon-lamp assembly L1 lights up, and transformer T1 changes 120 Vac to about 28 Vac. The rectifier bridge, D1 through D4, rectifies the ac into pulsating dc, which is then filtered by C1. Capacitor C1 acts as a storage capacitor. Zener diode D5 keeps the voltage constant across the base of Darlington regulator Q1, causing constant voltage across resistor R3 and the (+) and (–) output terminals, where the load is connected. Fuse F2 is used to open ("blow"), if the current through the output terminals is too high. Make sure to take proper precautions when using projects powered by 120 Vac.

0- TO 12-V, 1-A VARIABLE POWER SUPPLY



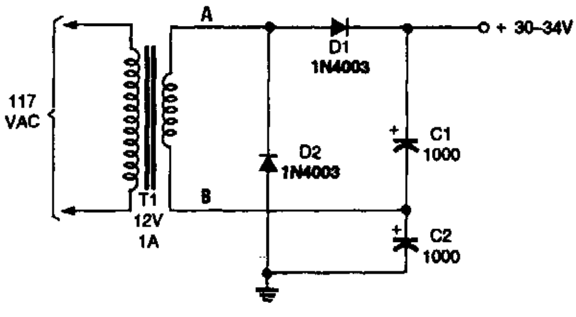
1991 PE HOBBYIST HANDBOOK ***FIG. 78-16***

This 0- to 12-Vdc variable power supply uses an IC voltage regulator and a heavy-duty transformer to provide a reliable dc power supply. Looking at the schematic shown, you can see that transformer T1 has a 120-V primary and a 28-V secondary.

Filtered dc is fed to the input (pin 2) of the LM317T voltage regulator, IC, which keeps the voltage at its output constant (pin 3) regardless (within limitations) of the input voltage. Pin 1 of the LM317T is the adjustment pin. Varying the voltage on pin 1 (via P1) varies the output voltage.

Diodes D5 through D7 and LEDs L1 through L3 give an approximate indication of the output voltage. Each LED/diode path has a limiting resistor to limit the current to a level that is safe for the LED.

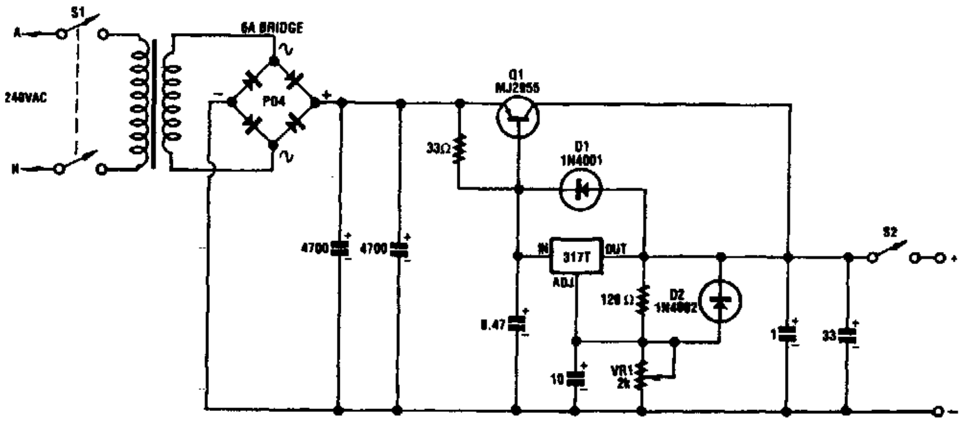
VOLTAGE DOUBLER SUPPLY



POPULAR ELECTRONICS ***FIG. 78-17***

The voltage doubler is built around a pair of diodes (D1 and D2) and a pair of capacitors (C1 and C2) that are fed from, in this case, a 12-V, 1-A step-down transformer (T1).

ADJUSTABLE 20-V SUPPLY



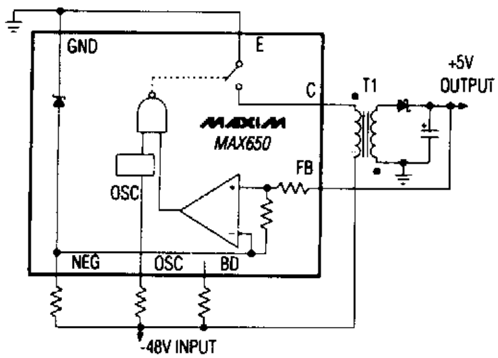
SILICON CHIP

FIG. 78-18

This circuit can deliver 3 A or more and a maximum dc voltage of a little over 20 V. It is designed around the readily available LM317T adjustable 3-terminal regulator and has a pnp power transistor to boost the current output.

The transformer has an 18-V secondary rated at 6 A; this feeds to bridge rectifier and two 4700-μF capacitors to yield around 25 Vdc. This voltage is fed to the emitter of the MJ2955 transistor and to the input of the LM317 via a 33-Ω resistor.

SWITCHING REGULATOR CONVERTER

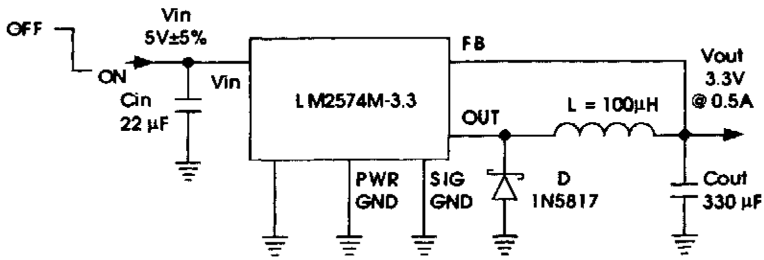


MAXIM ENGINEERING JOURNAL

FIG. 78-19

The Max650 switching regulator produces a regulated 5 V from large negative voltages, such as the –48 V found on telephone lines. The resulting power supply operates with several external components, including a transformer, and it delivers 250 mA. The device includes a 140-V 250-mA pnp transistor, short-circuit protection, and all necessary control circuitry.

5-V TO 3.3-V SWITCHING REGULATOR



NATIONAL SEMICONDUCTOR, LINEAR EDGE

FIG. 78-20

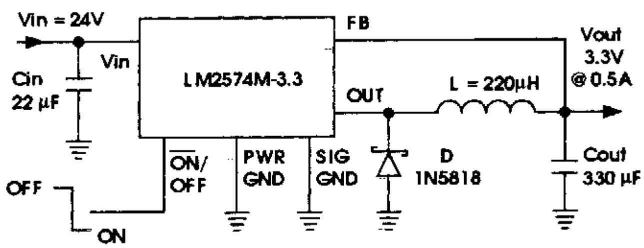
A National Semiconductor LM2574 is used to derive 3.3 V at 0.5 A from a 5-V logic bus. The duty cycle is:

$$\frac{V_{\mathrm{OUT}} + V_D - V_{IND}}{V_{\mathrm{IN}} - V_{SAT} + V_D - 2\,V_{IND}}$$

V_D = diode drop (0.39)
V_{IND} = inductor dc drop
V_{SAT} = saturation voltage of LM2574 (0.9 V typical)

This circuit should be useful to derive 3.3 V for logic devices from existing +5-V buses.

24-V TO 3.3-V SWITCHING REGULATOR



NATIONAL SEMICONDUCTOR, LINEAR EDGE

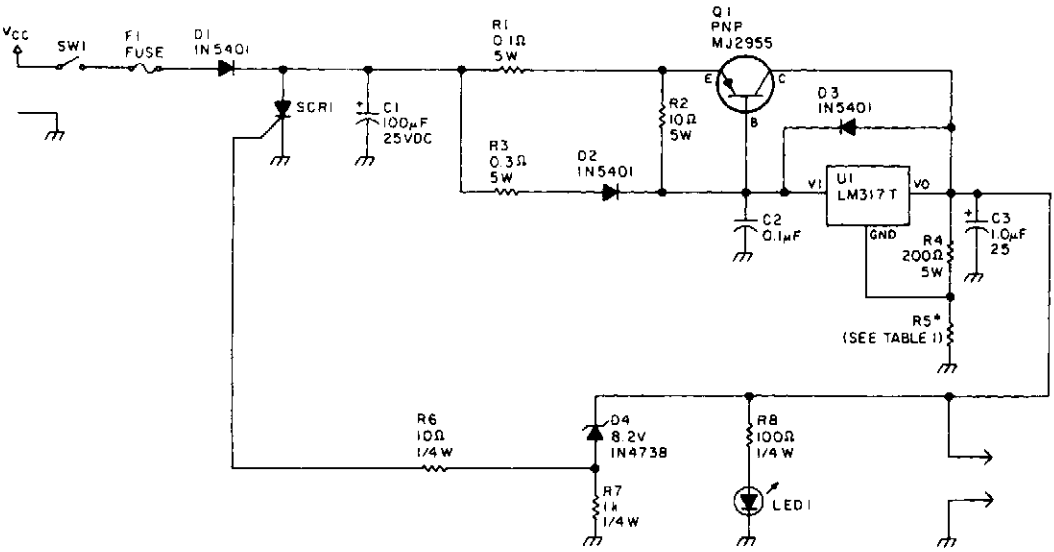
FIG. 78-21

The National Semiconductor LM2574 delivers 3.3 V out at 0.5 A from a 24-V source. The duty cycle is:

$$\frac{V_{\mathrm{OUT}} + V_D - V_{IND}}{V_{\mathrm{IN}} - V_{SAT} + V_D - 2\,V_{IND}}$$

V_D = diode drop (0.39)
V_{IND} = inductor dc drop
V_{SAT} = saturation voltage of LM2574 (0.9 V typical)

LAPTOP COMPUTER POWER SUPPLY



R5 Resistor Value	Voltage Out
750Ω	5V
910Ω	6V
1.2K	8V
1.5K	9V
1.8K	10V
2.0K	12V
2.7K	15V
3.3K	18V
3.6K	20V
4.3K	24V

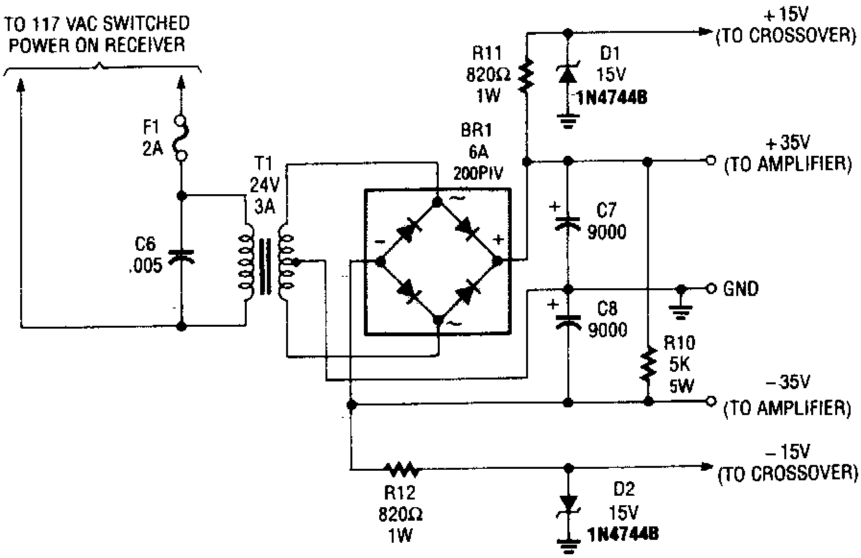
Note: Any output voltage value greater than 10V requires a higher input voltage than 13.6V. In addition capacitor working voltage ratings will have to be increased accordingly. Allow a minimum of 2.5 times the voltage expected to appear across the capacitor as a standard for the working voltage.

Table 1. Resistor value/voltage matchup.

73 AMATEUR RADIO TODAY **FIG. 78-22**

A laptop computer supply that has 9-V output, crowbar overvoltage protection, and operates from a 12-V supply is shown above. The supply voltage should be at least 3.6 V above the expected output voltage. Q1 should be heatsinked appropriately. R5 should have a value of 1.5 kΩ for 9-V output. Table 1 gives values for other voltages.

SUBWOOFER AMPLIFIER POWER SUPPLY

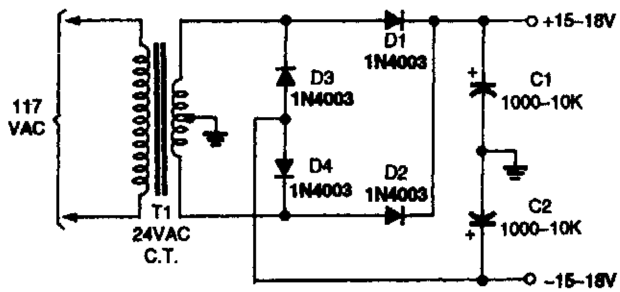


POPULAR ELECTRONICS

FIG. 78-23

Although intended to power a 100-W low-frequency amplifier, this power supply should handle many mono or stereo amplifiers in the medium power range that require ±30 to 35 V.

DUAL VOLTAGE-RECTIFIER CIRCUIT

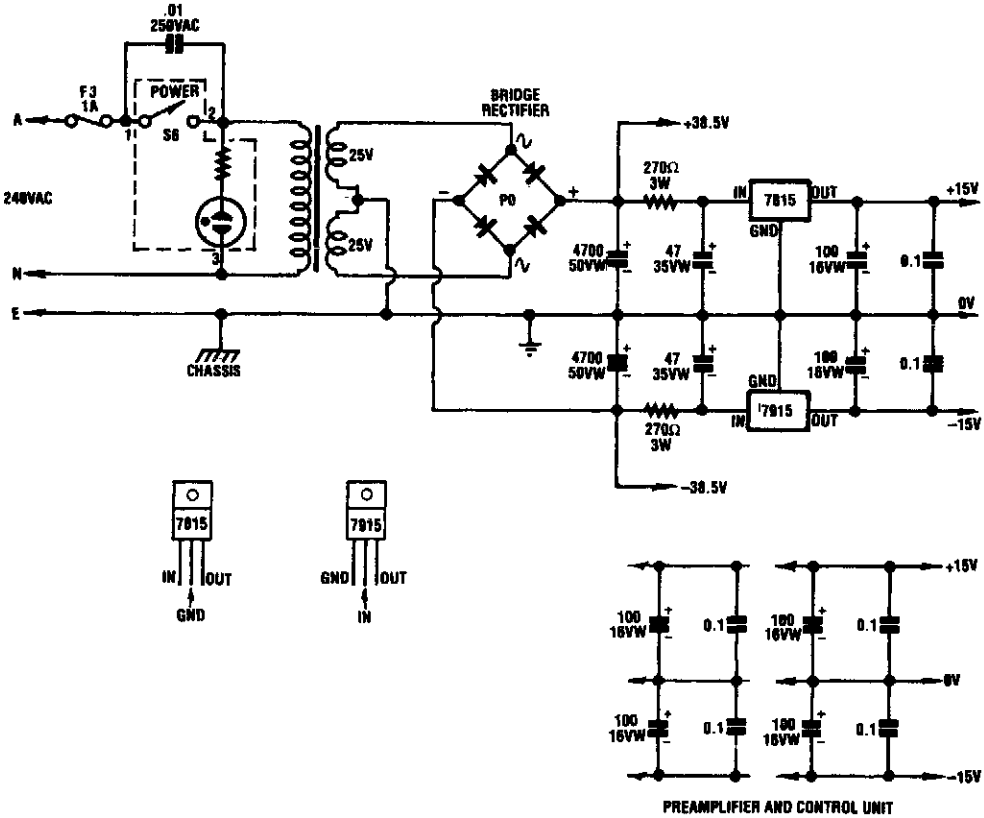


POPULAR ELECTRONICS

FIG. 78-24

This stepped-up dual voltage supply provides ±15 to ±18 V unregulated.

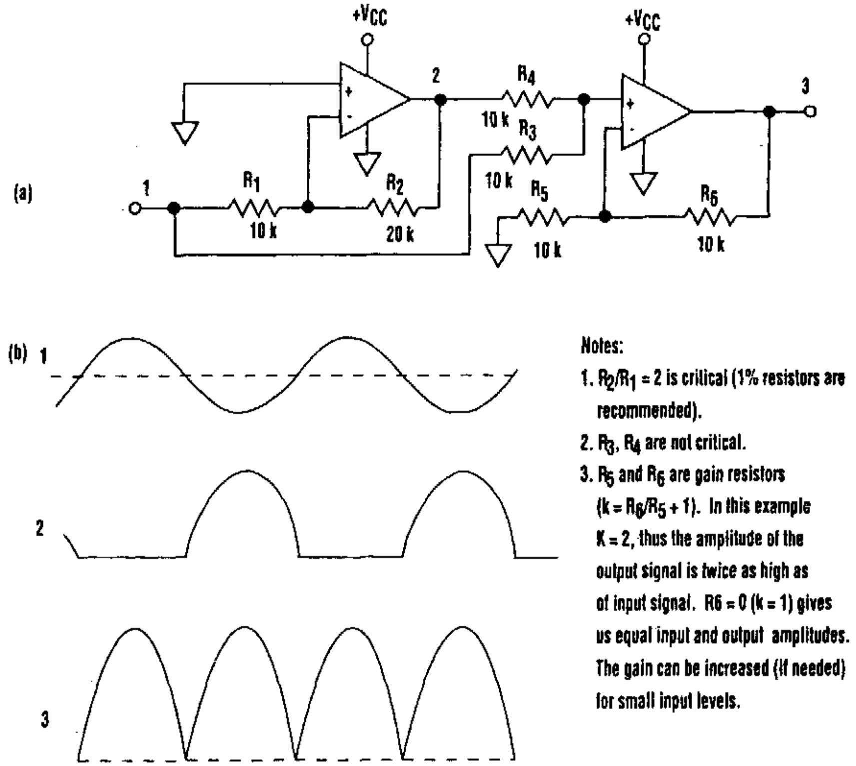
DUAL AUDIO AMPLIFIER POWER SUPPLY



SILICON CHIP *FIG. 78-25*

A dual audio amplifier that will deliver 50 W per channel is shown in the schematic. It includes preamp and tone controls, and also includes a headphone amplifier. The circuit depicts the power supply that supplies ±38.5 V and ±15 V regulated for the dual 50 watter.

DIODELESS RECTIFIER



ELECTRONIC DESIGN

FIG. 78-26

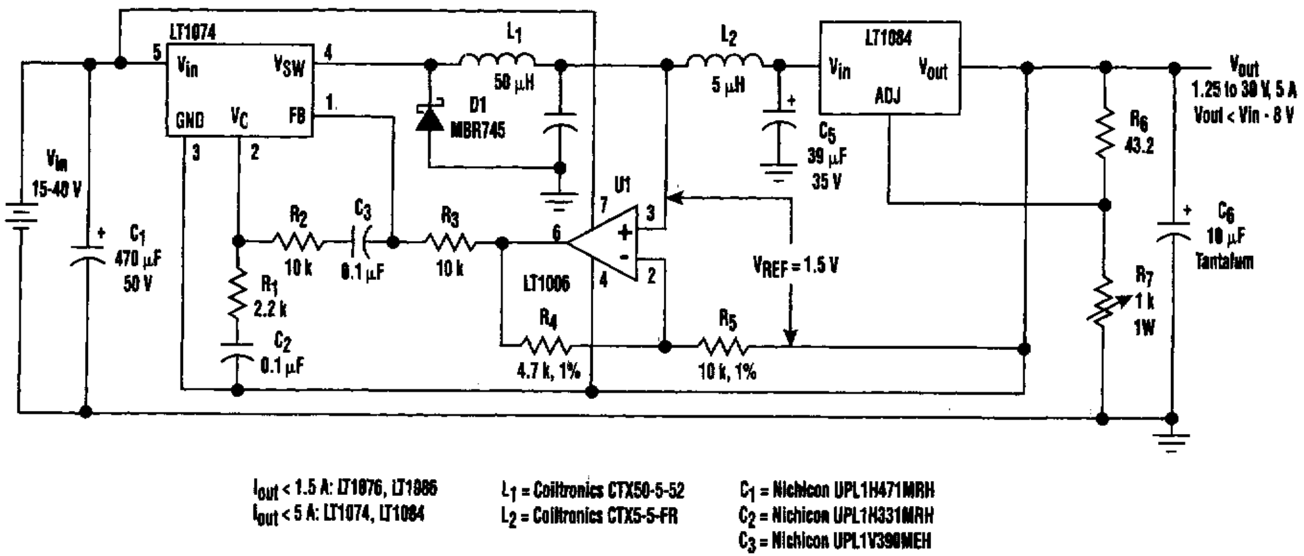
It's common knowledge that when working with single-supply op amps, implementing simple functions in a bipolar signal environment can be difficult. Sometimes additional op amps and other electronic components are required.

Taking that into consideration, can any advantage be attained from this mode? The answer lies in this simple circuit (A). Requiring no diodes, the circuit is a high-precision full-wave rectifier with a high-frequency limitation equalling that of the op amps themselves. Look at the circuit's timing diagram (B) to see the principle of operation.

The first amplifier rectifies negative input levels with an inverting gain of 2 and turns positive levels to zero. The second amp, a noninverting summing amplifier, adds the inverted negative signal from the first amplifier to the original input signal. The net result is the traditional waveform produced by full-wave rectification.

In spite of the limitation on the input signal amplitude (it must be less than $V_{CC}/2$), this circuit can be useful in a variety of setups.

REGULATOR LOSS CUTTER



ELECTRONIC DESIGN

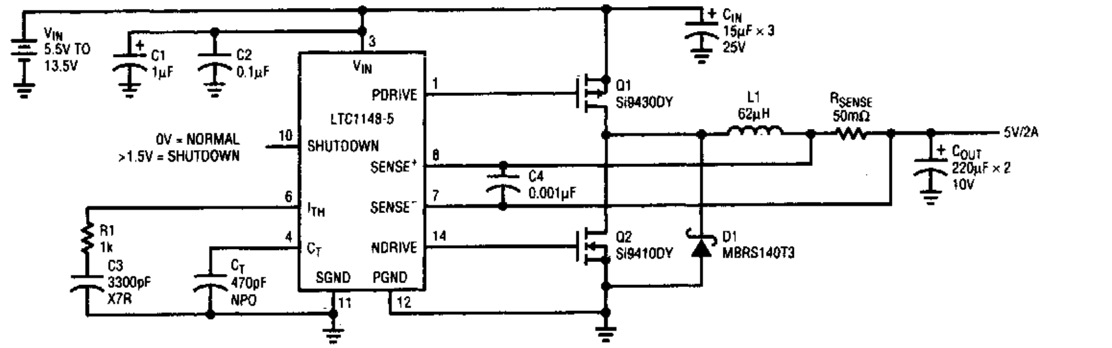
FIG. 78-27

Large input-to-output voltage differentials, caused by wide input voltage variations, reduce a linear regulator's efficiency and increase its power dissipation. A switching preregulator can reduce this power dissipation by minimizing the voltage drop across an adjustable linear regulator to a constant 1.5-V value.

The circuit operates the LT1084 at slightly above its dropout voltage. To minimize power dissipation, a low-dropout linear regulator was chosen. The LT1084 functions as a conventional adjustable linear regulator with an output voltage that can be varied from 1.25 to 30 V.

Without the preregulator (for a 40-V input and a 5-V output at 5 A), it would be virtually impossible to find a heatsink large enough to dissipate enough energy to keep the linear-regulator junction temperature below its maximum value. With the preregulator technique, however, the linear regulator will dissipate only 7.5 W under worst-case loading conditions for the entire input-voltage range of 15 to 40 V. Even under a short-circuit fault condition, the 1.5-V drop across the LT1084 is maintained.

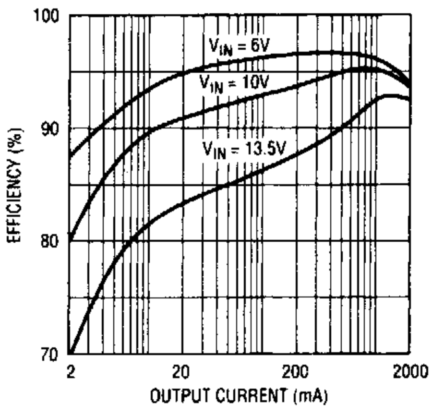
SYNCHRONOUS STEPDOWN SWITCHING REGULATOR WITH 90% EFFICIENCY



C1 (TA)
C_{IN} AVX (TA) TAJD156K025RLR, ESR = 0.3Ω, I_{RMS} = 0.707A
C_{OUT} AVX (TA) TAJE227K010RLR, ESR = 0.08Ω, I_{RMS} = 1.4A
Q1 SILICONIX PMOS, BVDSS = 20V, RDS_{ON} = 0.1Ω, C_{RSS} = 400pF, Q_G = 50nC
Q2 SILICONIX NMOS, BVDSS = 30V, RDS_{ON} = 0.05Ω, C_{RSS} = 160pF, Q_G = 30nC

D1 MOTOROLA SCHOTTKY, VBR = 40V
R_{SENSE} IRC LR2512-01-R050J P_D = 1W
L1 COILTRONICS CTX62-2-MP, DCR = 0.035Ω, MPP CORE (THROUGH HOLE)
L1-1 COILTRONICS CTX02-11715-2, DCR = 0.11Ω, FERRITE CORE (SURFACE MOUNT)
ALL OTHER CAPACITORS ARE CERAMIC

A LTC1148 (5.5V-13.5V to 5V/2A) surface mount

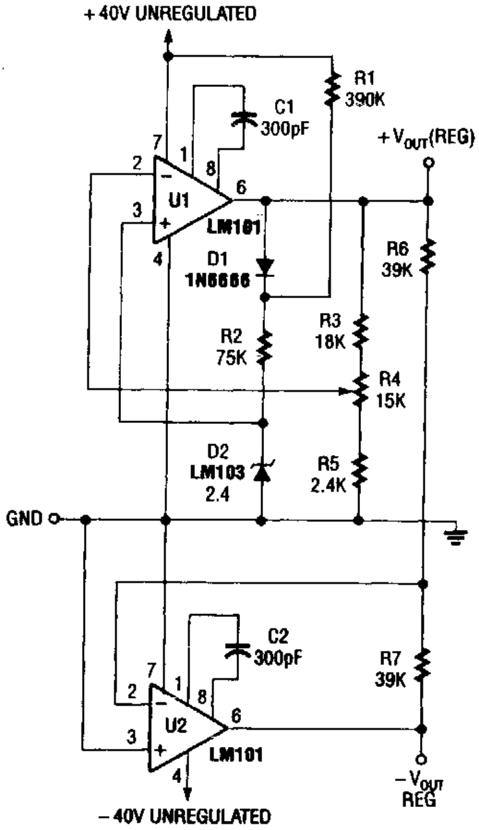


B LTC 1148-5: 5.5V to 13.5V efficiency

LINEAR TECHNOLOGY ***FIG. 78-28***

A shows a typical LTC1148 surface-mount application providing 5 V at 2 A from an input voltage of 5.5 V to 13.5 V. The operating efficiency, shown in B, peaks at 97% and exceeds 90% from 10 mA to 2 A with a 10-V input. Q1 and Q2 comprise the main switch and synchronous switch, respectively, and inductor current is measured via the voltage drop across the current shunt. R_{SENSE} is the key component used to set the output current capability according to the formula $I_{OUT} = 100\ mV/R_{SENSE}$. The advantages of current control include excellent line and load transient rejection, inherent short-circuit protection and controlled startup currents. Peak inductor current is limited to 150 mV/R_{SENSE} or 3 A for the circuit in A.

±5- TO ±35-V TRACKING POWER SUPPLY

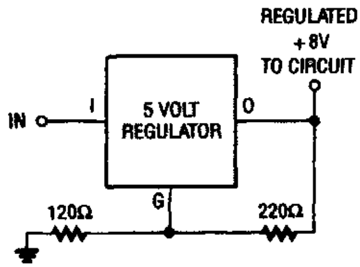


OUTPUT VOLTAGE IS VARIABLE FROM ±5V TO ±35V.
NEGATIVE OUTPUT TRACKS POSITIVE OUTPUT TO WITHIN THE RATIO OF R6 TO R7.

POPULAR ELECTRONICS ***FIG. 78-29***

This supply is designed to operate from a ±40-V nominal unregulated power source (bridge rectifier, etc.).

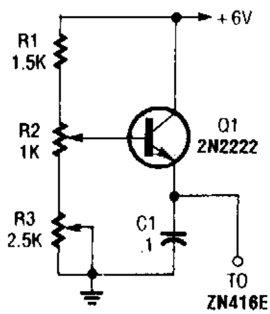
8-V FROM 5-V REGULATOR



POPULAR ELECTRONICS ***FIG. 78-30***

If you have trouble locating an 8-V regulator, although they are commonly available, a 5-V unit can replace it by connecting the regulator, as is shown here.

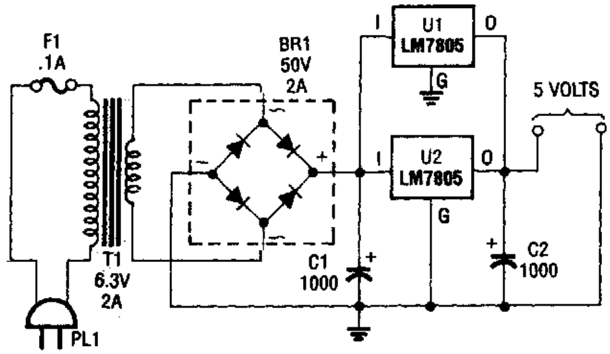
+1.5-V SUPPLY FOR ZN416E CIRCUITS



POPULAR ELECTRONICS ***FIG. 78-31***

This regulator can be used with a +6-V source to supply ZN416E low-voltage TRF radio-receiver IC the necessary +1.5 V. R3 sets output voltage.

ANTIQUE RADIO dc FILAMENT SUPPLY

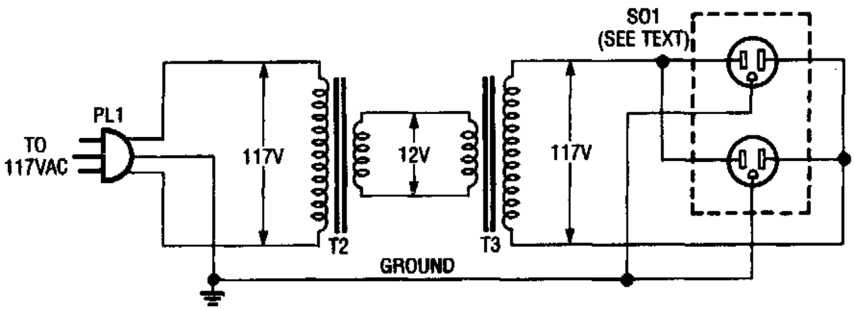


POPULAR ELECTRONICS

FIG. 78-32

This dc supply is great for operating battery-powered antique radios, because it is designed to prevent harming the tube filaments. The circuit is useful for powering filaments of 00-A, 01-A, 112A, and 71A tubes, which require 5V at 250 mA.

INEXPENSIVE ISOLATION TRANSFORMER (IMPROMPTU SETUP)

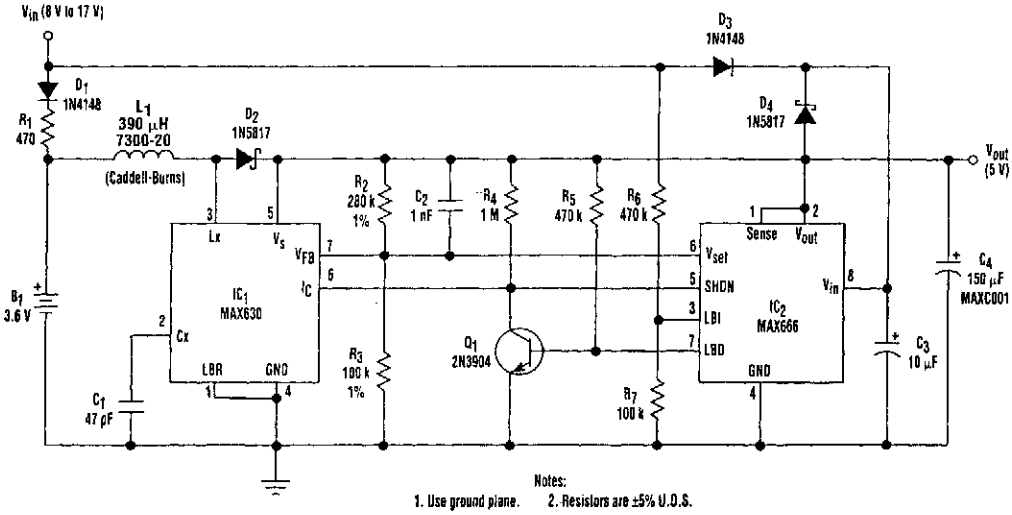


1993 ELECTRONICS HOBBYISTS HANDBOOK

FIG. 78-33

Using two 12-V filament or power transformers, an impromptu isolation transformer can be made for low-power (under 50 W) use in testing or servicing. SO1 is an ordinary, duplex ac receptable. Use heavy-wire connections between the 12-V windings because several amperes can flow.

5-V UPS

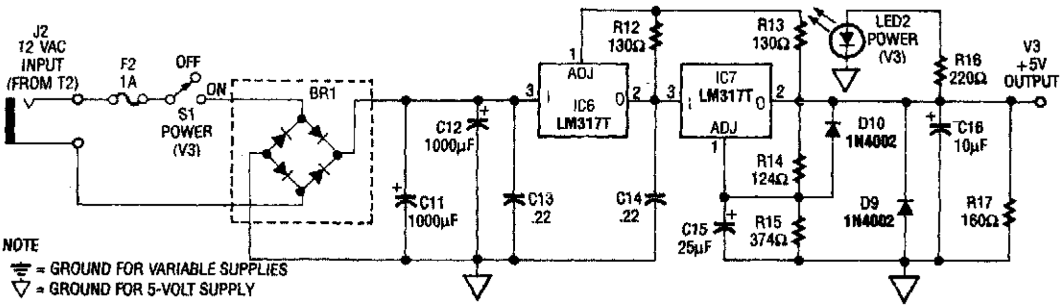


ELECTRONIC DESIGN *FIG. 78-34*

A 9-V wall adapter supplies V_{IN}. IC2 contains a low-battery detector circuit that senses V_{IN} by means of R6 and R7. The detector output (pin 7) drives an inverter (Q1), which in turn drives the shut-down inputs I_C of IC1 and SHDN of IC2. These inputs have opposite-polarity active levels. The common feedback resistors, R2 and R3 enable both regulators to sense the output voltage, V_{OUT}.

When IC2 shuts down, its output turns off. However, when IC1 shuts down, the whole chip assumes a low-power state and draws under 1 μA. L1, D2, C1, C2, R2, and R3 are part of the 250-mW switching regulator. Diodes D3 and D4 wire-OR the power connection to IC2, and C3 improves the linear regulator's load regulation.

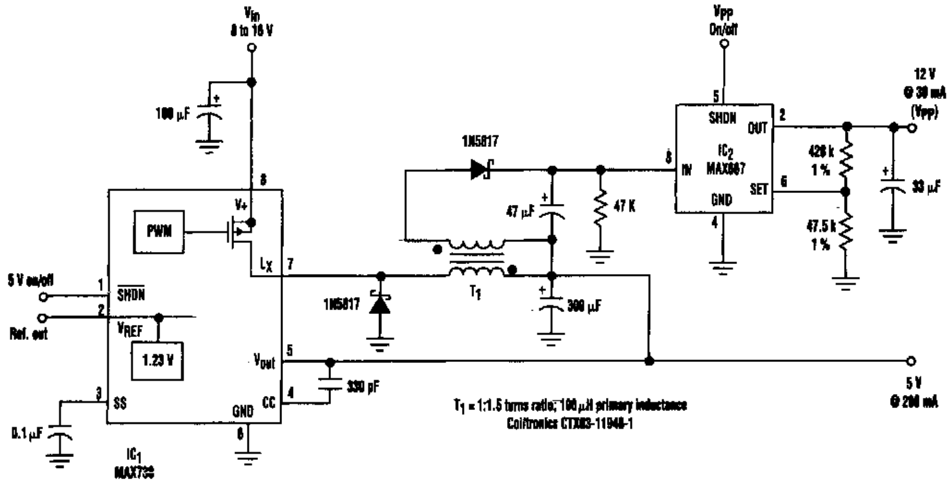
+5-V SUPPLY



ELECTRONICS NOW *FIG. 78-35*

The power supply shown is designed to operate from a wall transformer. This circuit can be used in conjunction with a variable supply to test circuits in the lab, etc. T2 is a 12-V wall transformer.

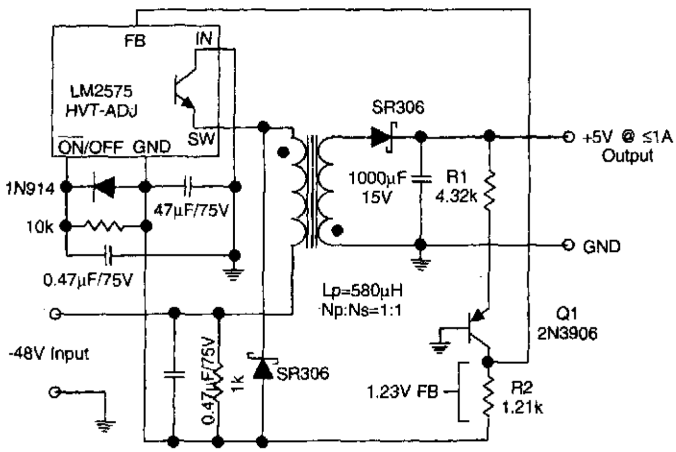
ADD 12-V OUTPUT TO 5-V BUCK REGULATOR



ELECTRONIC DESIGN ***FIG. 78-36***

By adding a flyback winding to a buck-regulator switching converter (see the figure), which is essentially a 5-V supply with a 200-mA output capability, a 12-V output (V_{pp}) can be produced. The flyback winding on the main inductor (forming transformer T1) enables an additional low-dropout linear regulator (IC2) to create the 12-V output voltage that's needed to program EEPROMs. The required input voltage is 8 to 16 V.

TELECOM CONVERTER –48 V TO +5 V @ 1 A



NATIONAL SEMICONDUCTOR, LINEAR EDGE ***FIG. 78-37***

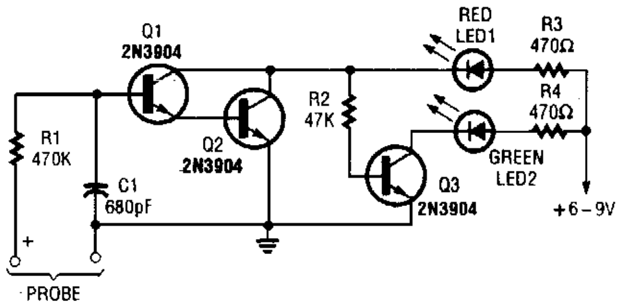
The circuit supplies 1 A at +5 V from the –48-V supply commonly used in telephone equipment. The National Semiconductor LM2575 is a simple switching regulator.

79

Probe Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

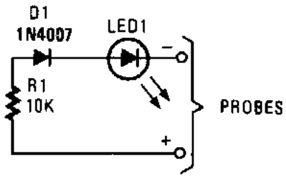
SIMPLE VOLTAGE PROBE



POPULAR ELECTRONICS ***FIG. 79-1***

This simple voltage probe can be helpful in checking and troubleshooting solid-state circuitry.

ac VOLTAGE PROBE



POPULAR ELECTRONICS ***FIG. 79-2***

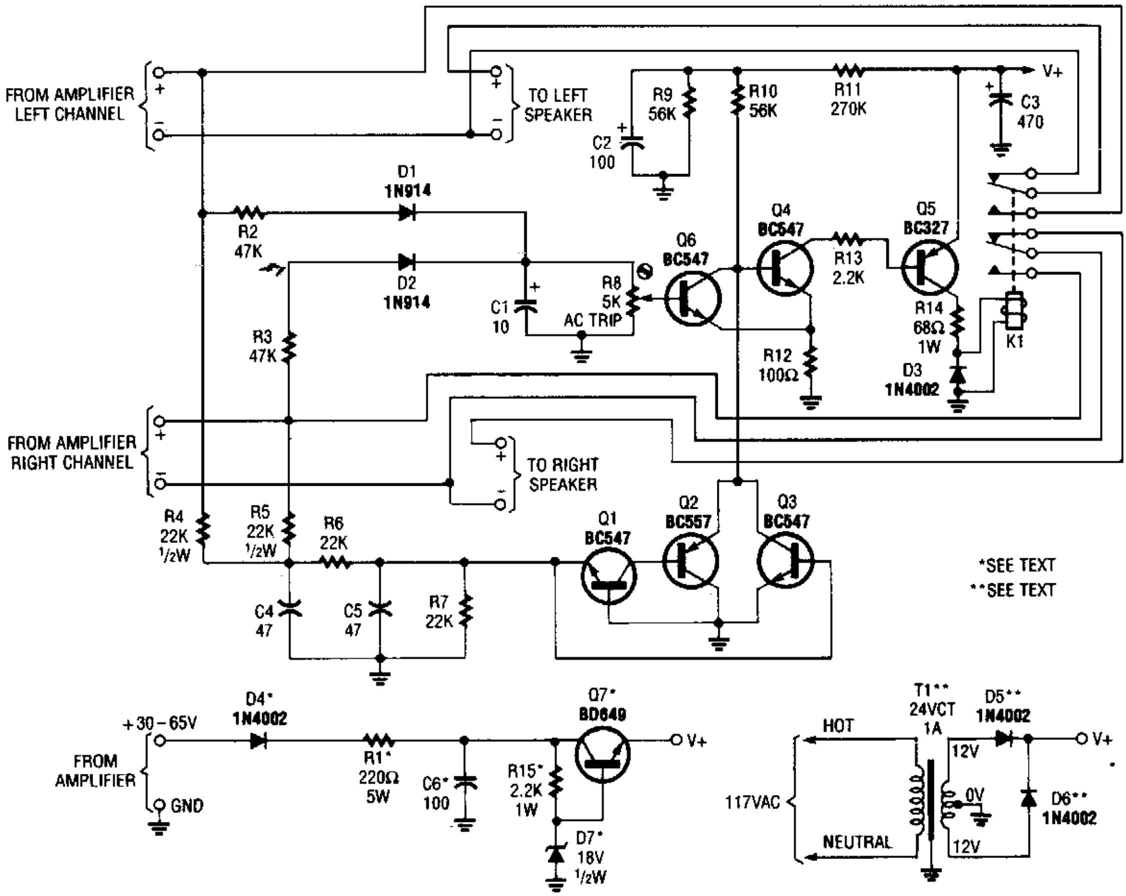
This simple probe can save your life by warning you of live circuitry. It's ideal for times when more than one person is working on a device.

80

Protection Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

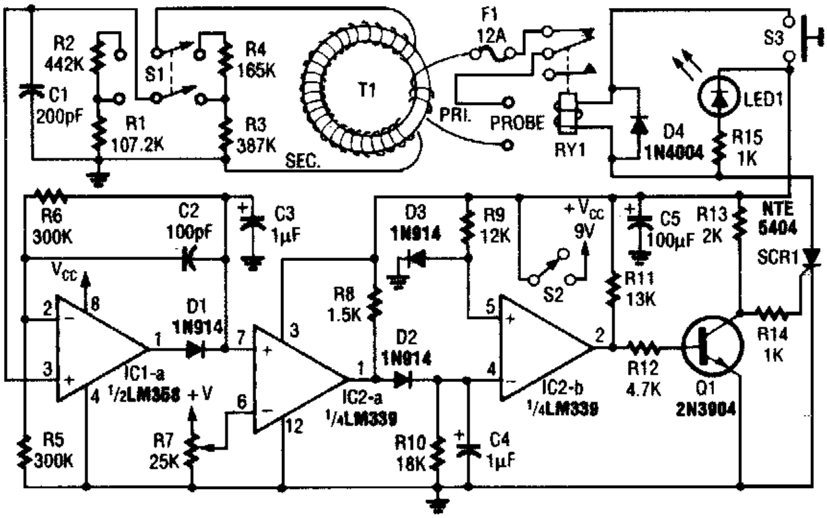
SPEAKER PROTECTOR



POPULAR ELECTRONICS *FIG. 80-1*

Most of the transistors in this speaker protector function as switches. Normally, Q4, Q5, and K1 are on and the speakers are connected to the amplifier. However, if a large dc voltage appears at an amplifier output, either Q3, or Q1 and Q2 turn on, biasing Q4 off. That action turns Q5 off, de-energizes the relay, and disconnects the speakers from the amplifier. Components D1, D2, and Q6 form the overdrive-protection circuit.

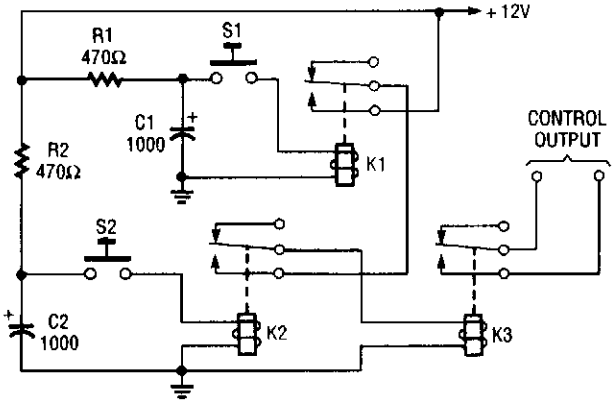
ELECTRONIC FUSE



RADIO-ELECTRONICS *FIG. 80-2*

Basically, this circuit is an adjustable electronic circuit breaker, containing a toroidal transformer that senses 60-Hz load current. T1 has a two-turn winding for primary, and 100 turns of #30 gauge wire for the secondary. A high-low range switch selects 0.1 to 6 A or 1 to 12 A. The primary winding of T1 carries full load current and voltage; should be suitably insulated, as should be RY1.

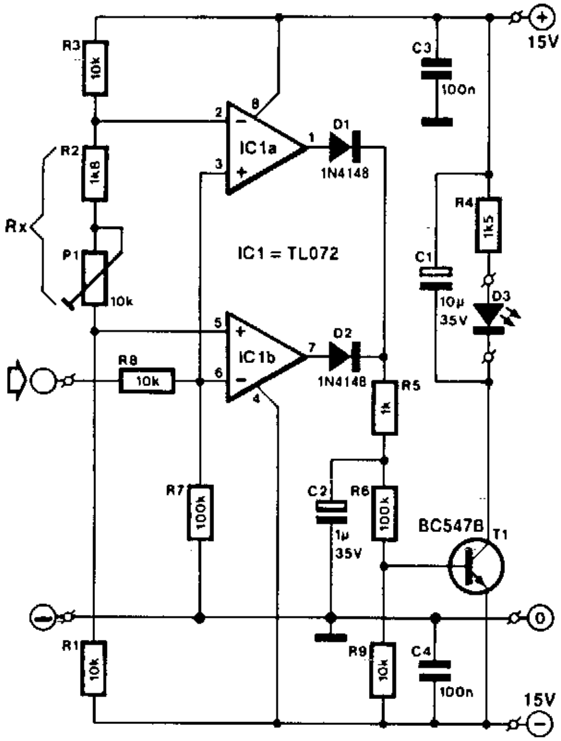
SAFETY CIRCUIT



POPULAR ELECTRONICS *FIG. 80-3*

Because of the finite hold-on time of delay circuits R1/C1 and R2/C2, both S1 and S2 must be pressed at the same time to power up the load.

OVERLOAD INDICATOR

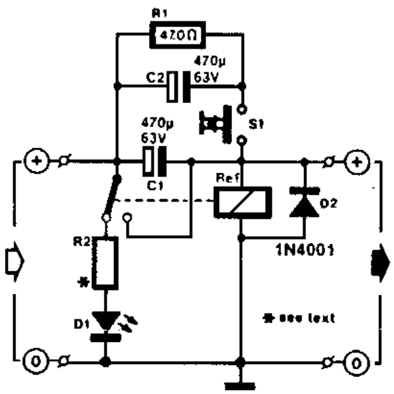


Two op amps are used as comparators to indicate excessive magnitude of an AF signal, either positive or negative, even if the signal is asymmetrical. P1 sets the reference voltage for both op amps. This circuit is useful for audio-amplifier and op-amp circuits using split power supplies.

ELEKTOR ELECTRONICS

FIG. 80-4

RELAY FUSE FOR POWER SUPPLIES

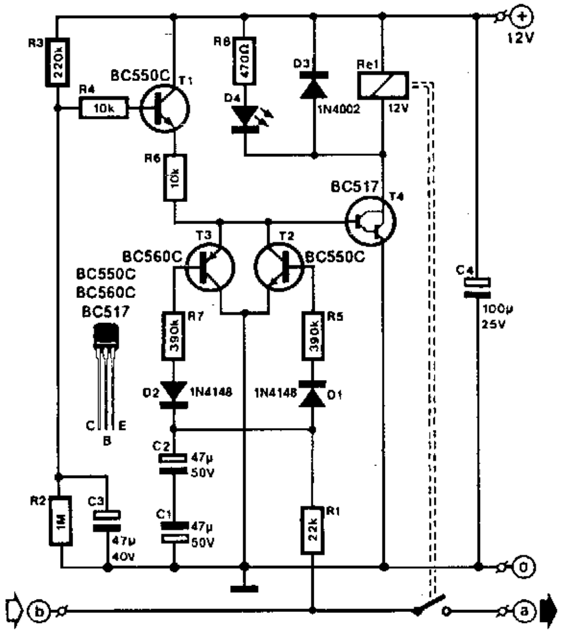


A method of adding overload protection to a power supply using a relay is shown. In each circuit, the relay must be reset by a momentary switch using a charge on capacitor C2. This prevents overload if the short still exists.

ELEKTOR ELECTRONICS

FIG. 80-5

SPEAKER PROTECTOR

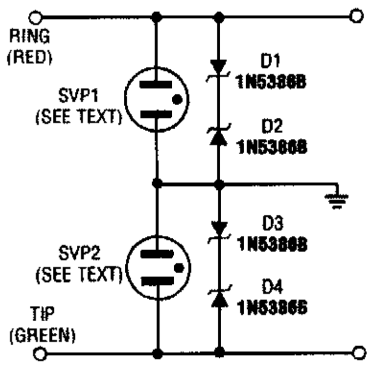


ELEKTOR ELECTRONICS

FIG. 80-6

A speaker system can be protected against amplifier failure when dc voltages (on speaker line a-b) are sensed by the circuit. Either positive or negative dc voltages are sensed. A relay opens in this case, removing the dc from the speakers. About 12 V at 50 mA is needed to power the circuit, depending on the relay.

MODEM PROTECTOR

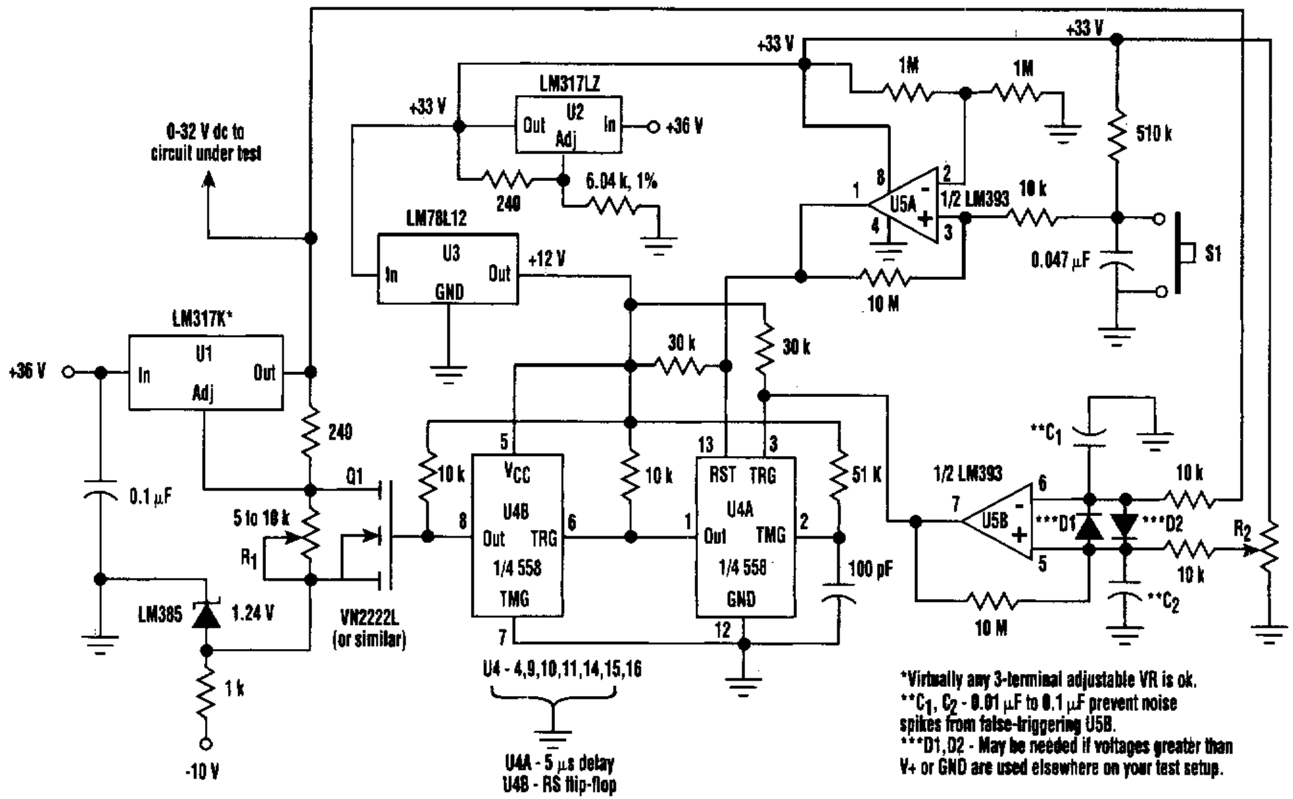


ELECTRONICS NOW

FIG. 80-7

This protector uses surge voltage protectors rated at 230-V breakdown. An effective ground should be used.

OVERVOLTAGE PROTECTION CIRCUIT

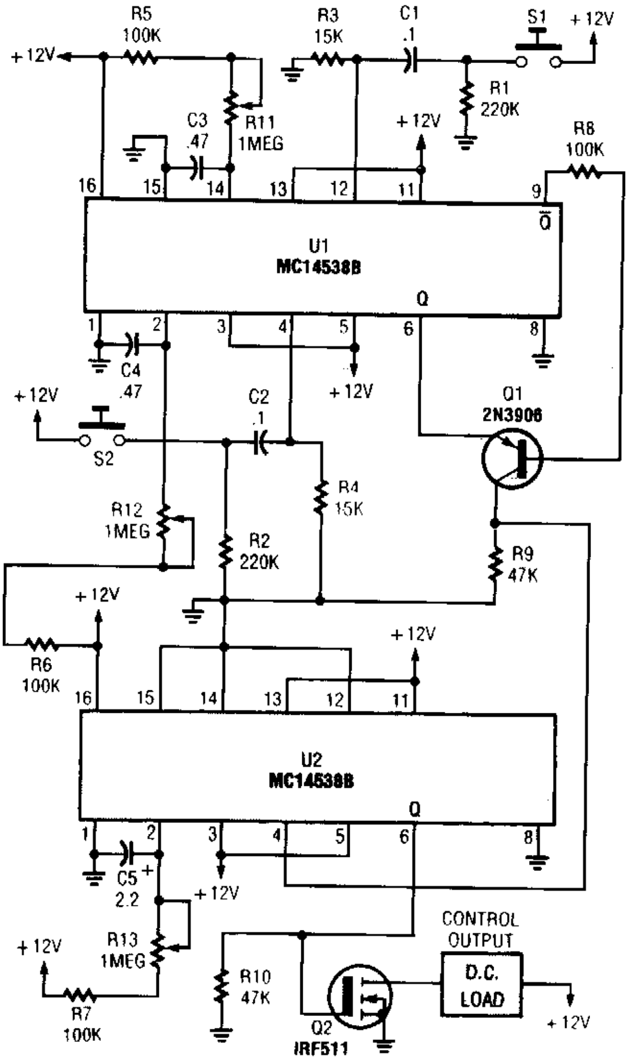


ELECTRONIC DESIGN

FIG. 80-8

When testing a circuit, a source of voltage that is variable and has overvoltage shutdown is very useful. In this circuit, R1 is adjusted to 1 to 2 V below the eventual shutdown threshold. R2 sets the trip voltage. When this voltage is reached, the circuit shuts the voltage to the circuit under test down. To reset, reduce R1 below trip threshold and depress reset switch S1.

TIMED SAFETY CIRCUIT

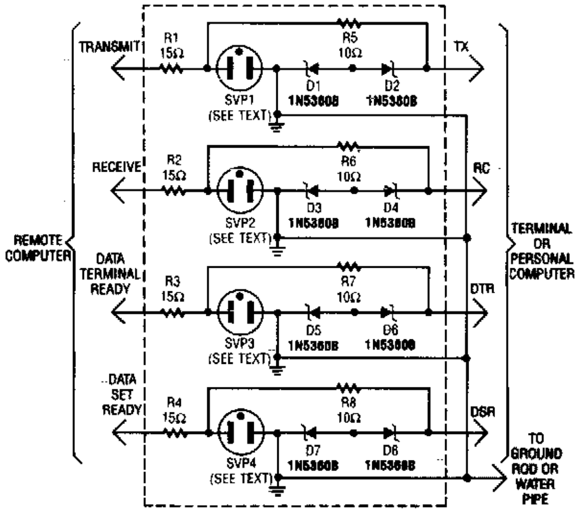


POPULAR ELECTRONICS

FIG. 80-9

When S1 is closed, pin 9 of U1 goes low, turning on Q1 for a preset period. If S2 is closed during this period, Q2 is turned on for a preset period. R11 and R13 set the two time periods.

MODEM /FAX PROTECTOR FOR TWO COMPUTERS

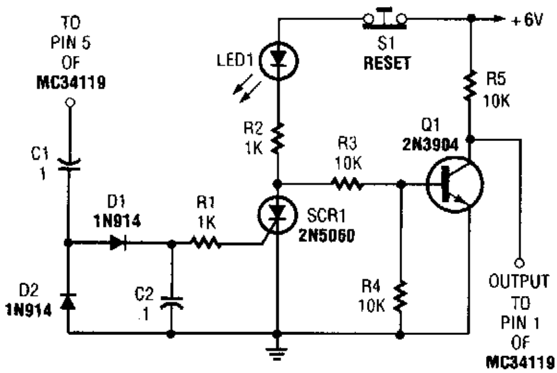


VARIATION OF THE MODEM/FAX PROTECTOR for use in telephone line connections between PC or terminal and larger distant computer.

ELECTRONICS NOW **FIG. 80-10**

This modem/fax protector can be used in telephone-line connections between a PC or a terminal and a distant computer. In this circuit, the SVPs (surge voltage protectors) are rated at 230 V. A good ground is a must for effective operation.

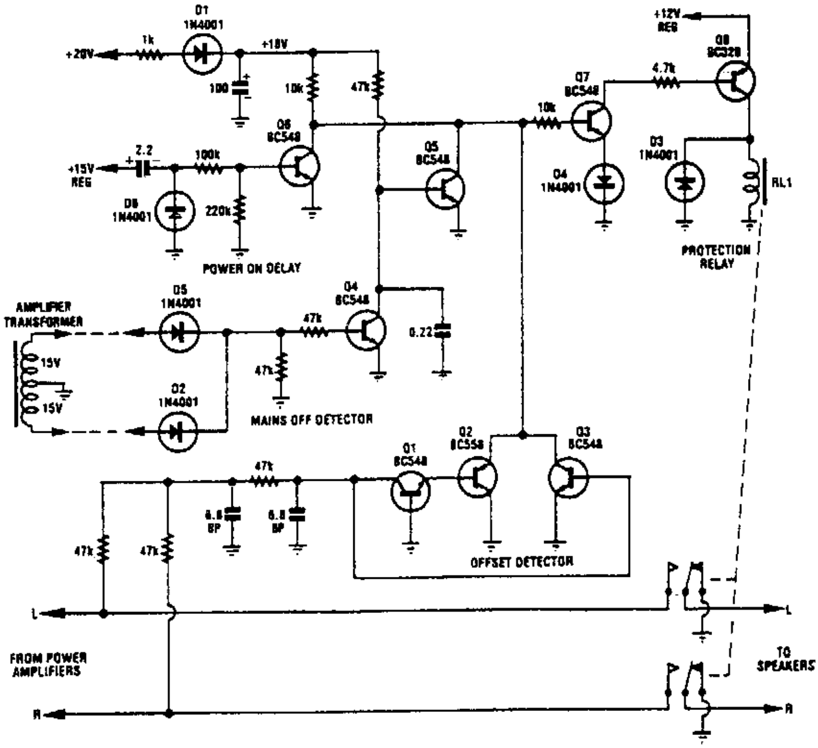
EAR PROTECTOR



POPULAR ELECTRONICS **FIG. 80-11**

The ear protector is actually a peak audio-detector/shutdown circuit that disables the amplifier through its chip-disable input when the output volume of an amplifier reaches the set level. The circuit, although intended for the MC34119 amplifier, should work with similar IC devices or applications.

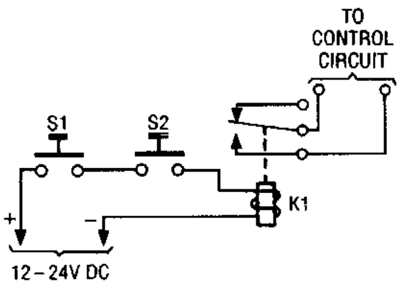
LOUDSPEAKER PROTECTOR



SILICON CHIP *FIG. 80-12*

Transistors Q1, Q2, and Q3 monitor the two outputs of the stereo amplifier. If the offsets exceed ±2 V, Q7 is turned off, which turns off Q8 and the normally on relay. Diodes D2 and D5, together with Q4, provide a mains voltage monitor. As soon as the ac input voltage disappears, as when the amplifier is turned off, Q4 turns off and Q5 turns on. This turns off Q7, Q8, and the relay. Hence, the loudspeakers are disconnected immediately after the amplifier is turned off.

SIMPLE SAFETY CIRCUIT



The simple two-hand safety-control switch shown here is little more than two pushbutton switches connected in series; both must be depressed in order to energize the relay.

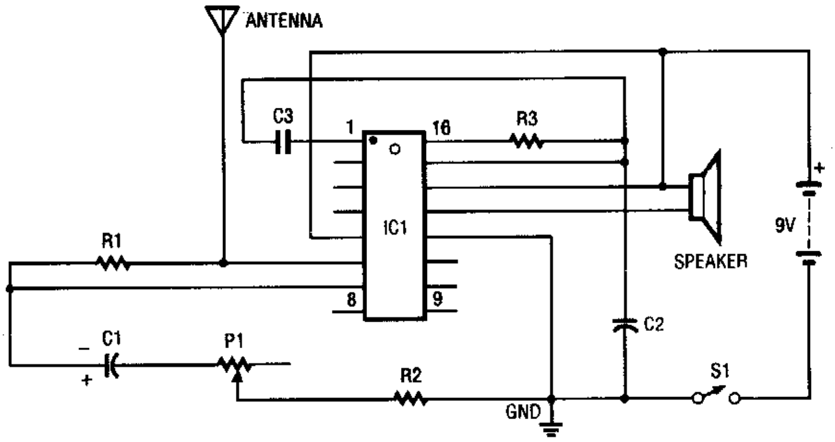
POPULAR ELECTRONICS *FIG. 80-13*

81

Proximity Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

PROXIMITY ALARM I



1991 PE HOBBYIST HANDBOOK

FIG. 81-1

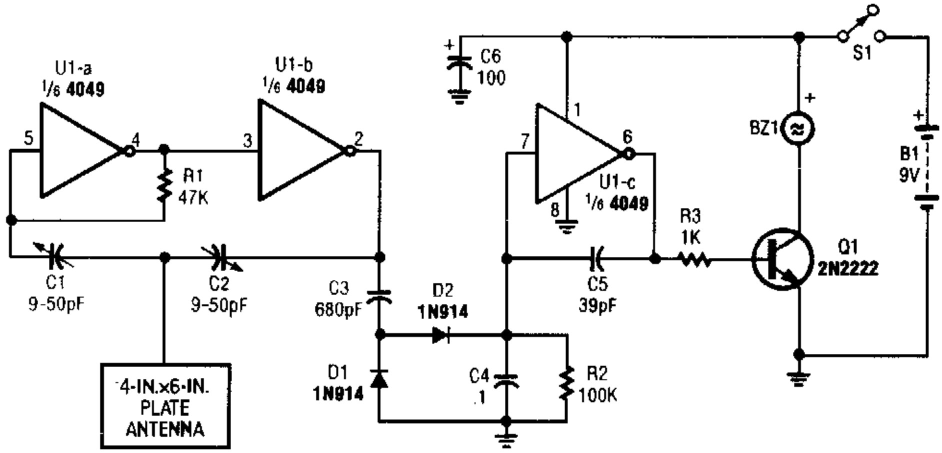
IC1 contains several oscillators and an amplifier. The low-frequency audio-signal oscillator is used to supply an input to the amplifier. That signal is the audio tone that is amplified, then supplied to the speaker by the amplifier.

The high-frequency oscillator is purposely set to be very unstable. It is dormant or "off" until the resistor-capacitor (RC) network is changed. The resistance (R) in this case is made up of R2 and P1. As the resistance of P1 is decreased, the unit becomes more sensitive (more unstable), and less capacitance (C) is needed to cause the oscillator to oscillate.

The capacitance required is provided by C2 and by any capacitance introduced via the antenna loop. When you come near that loop, your inherent body capacitance causes the high-frequency oscillator to begin to oscillate, which then causes the low-frequency oscillator to be "switched on" internally. Once the alarm is sounding, the IC is designed so that it "latches", that is, it stays on until the power to it is switched off.

C1	1-µF Axial Capacitor
C2	27-pF Silver Mica Capacitor
C3	0.1-µF Mylar Capacitor
IC1	CM1001N IC
P1	50-kΩ Trimmer Resistor
R1	75-kΩ Resistor
R2	200-Ω Resistor
R3	100-kΩ Resistor
S1	SPDT Switch
Spk	Small Speaker
Misc	IC Socket, Battery Snap, Ground Plate, Wire, PC Board

PROXIMITY ALARM II



POPULAR ELECTRONICS ***FIG. 81-2***

A CMOS logic gate is used to make up this circuit. When an object is near the antenna, the change in oscillator output is detected by D1 and D2 and amplified by U1C, which drives Q1, sounding alarm BZ1.

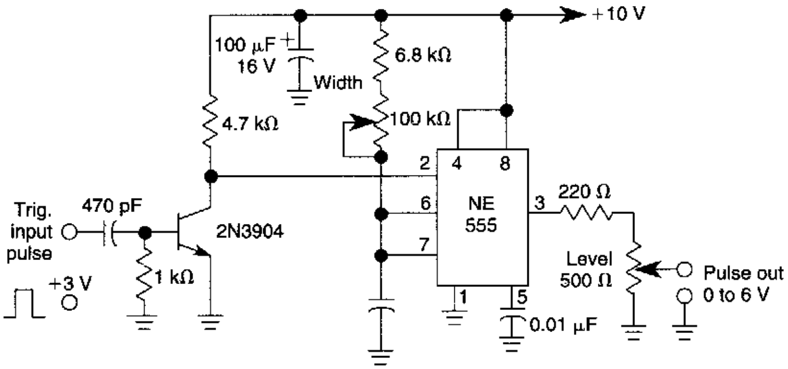
82

Pulse-Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Add-On Pulse Generator
Pulse Generator
Logic Pulser
Precise One-Shot
Digitally Controlled Sawtooth Pulse Generator
Delayed Pulse Generator
Pulse Generator with Variable Duty Cycle

ADD-ON PULSE GENERATOR



C μF	Pulse width
4.7	40 ms to 540 ms
0.47	4 ms to 54 ms
0.047	400 ms to 5.4 ns
0.0047	40 μs to 540 μs
470 pF	4 μs to 54 μs

WILLIAM SHEETS ***FIG. 82-1***

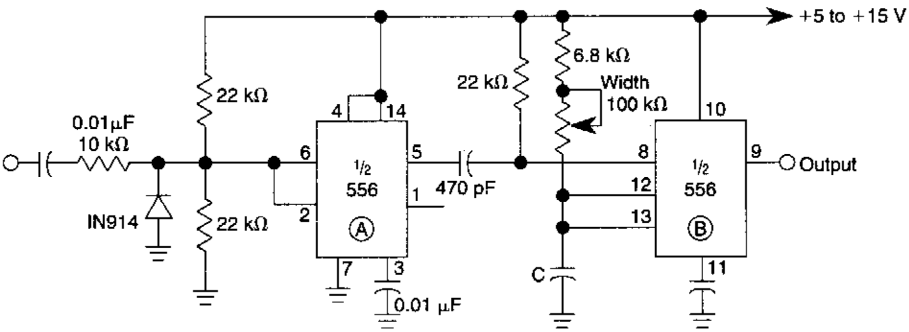
This pulse generator can supplement a standalone pulse generator. Using a transistor and a 555 timer, pulse widths of <5 μs to 500 μs can be produced. The value of C_3 is approximately found from the formula:

$C_3\ \mu F \approx 1.1 \times 10^{-5}\ T$ where T is the shortest pulse width (μs) desired in a 10:1 range

(T should be greater than 5 μs)

The capacitor values and consequent pulse width range are shown.

PULSE GENERATOR



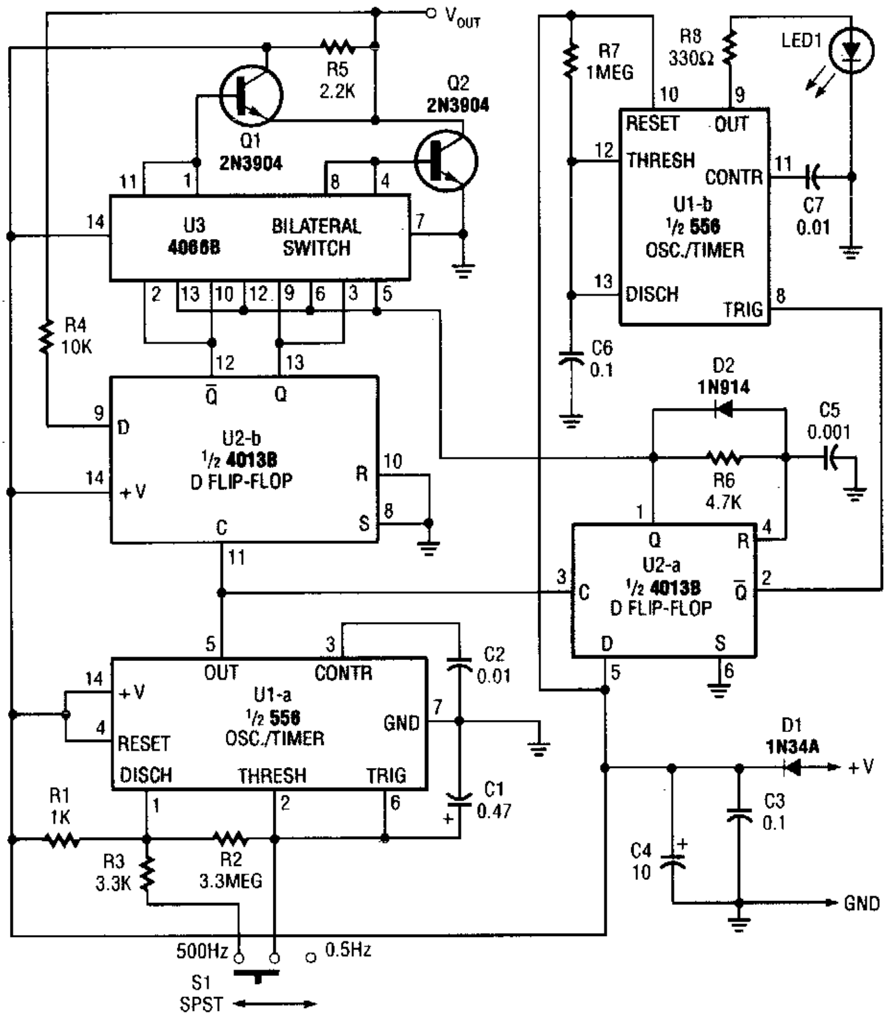
Pulsewidth $T \approx 1.1\ RC$

In this circuit $T \approx 7.4 \times 10^{-3}\ C_{\mu F}$ to $0.117 \times C_{\mu F}$ seconds
with $C = 0.1\mu F$ $T = 740\ \mu s$ to 11.7 ms

WILLIAM SHEETS ***FIG. 82-2***

By using a 556 dual timer with IC1A acting as a waveshaper and IC1B as a pulse generator, a 10:1 range of pulse widths can be generated.

A sine wave can be used to trigger this circuit.

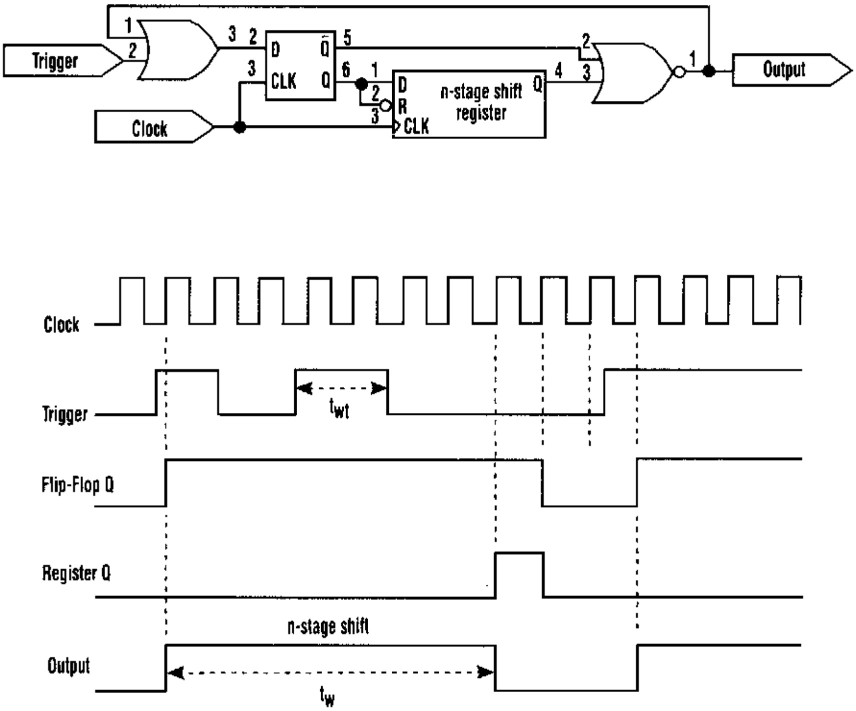


1990 PE HOBBYIST HANDBOOK

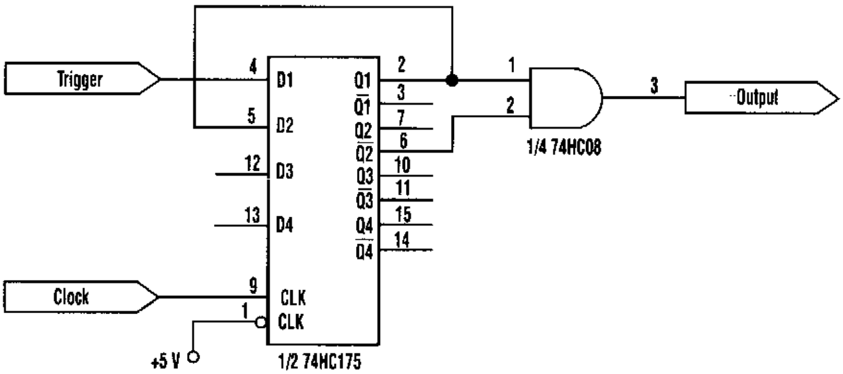
FIG. 82-3

The logic pulser generates pulses at 500 Hz or 0.5 Hz. When the pulser's tip connects to an input that is already being driven high or low, the pulser senses the logic state and automatically pulses the input briefly to the opposite state.

PRECISE ONE-SHOT



A more precise and stable one-shot pulse is generated by this circuit (a). When a trigger pulse is present, the flip-flop initiates a one-shot pulse whose width is a multiple of the clock period (b).



This simple one-shot circuit has a pulse width of one clock period and is more precise and stable than a multivibrator.

ELECTRONIC DESIGN ***FIG. 82-4***

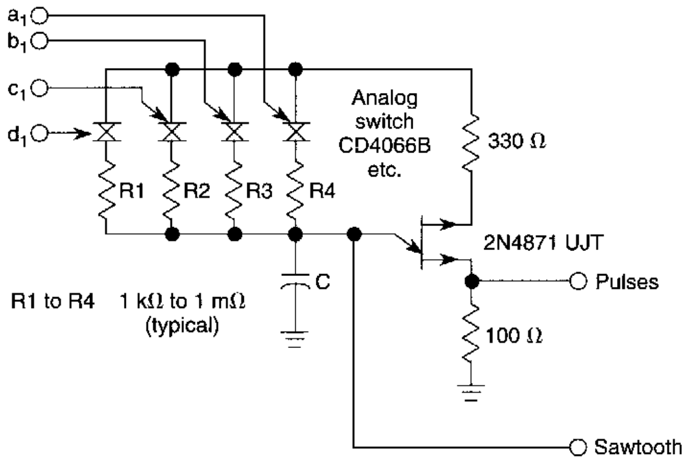
This approach uses a flip-flop, a shift register, and two gates (A). Before the one-shot pulse, the output of the NOR gate is 0. Consequently, the data input of the D-type flip-flop is equivalent to the trigger. When a trigger pulse is present, the flip-flop initiates the one-shot pulse, and the n-stage shift register controls the pulse width, t_w, which is a multiple of the clock's period (B).

The precision of the one-shot pulse is determined by the clock period, which is inversely proportional to its frequency. For the circuit to work properly, the width of the trigger pulse, t_{wt}, should be greater than one clock period.

The OR gate masks the trigger's effect when the circuit is generating the desired pulse. The net result is a circuit that functions as a nonretriggerable multivibrator.

When the pulse needs to be only one-clock-period wide, the circuit can be simplified. All that's required are two D-type flip-flops and an AND gate. However, despite its simplicity, this circuit generates a more stable and precise one-shot pulse than a multivibrator.

DIGITALLY CONTROLLED SAWTOOTH PULSE GENERATOR

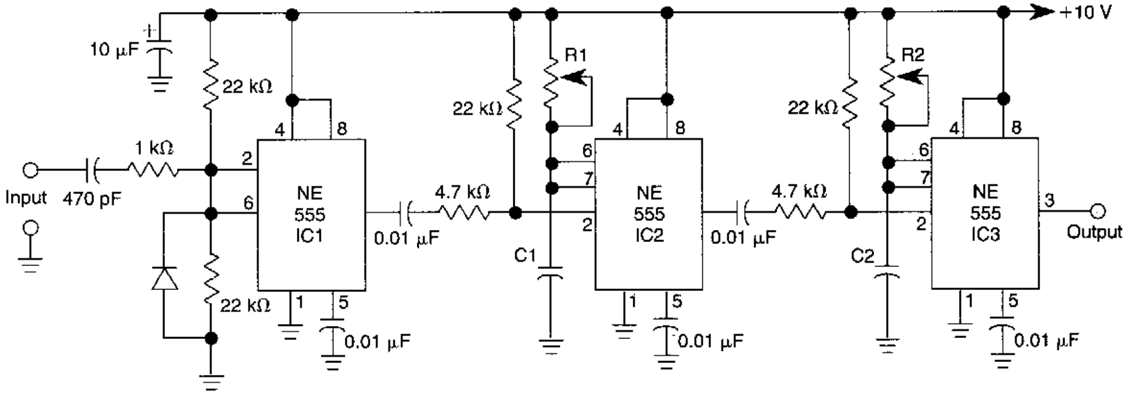


WILLIAM SHEETS

FIG. 82-5

Use of an analog switch as shown allows digital control of a UJT oscillator.

DELAYED PULSE GENERATOR



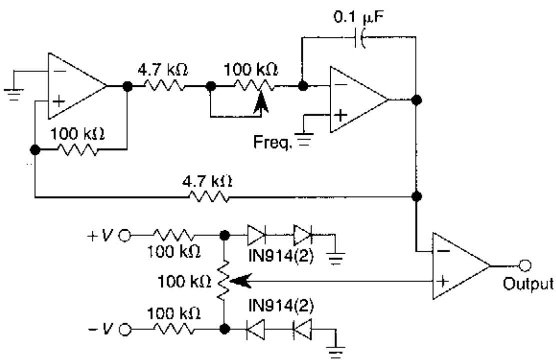
WILLIAM SHEETS

FIG. 82-6

Three 555 IC timers are used in this circuit to construct a simple delayed-pulse generator. IC1 acts as a waveform shaper to produce a rectangular waveform. IC2 produces a delaying pulse to trigger IC3 on the trailing edge of the delaying pulse. R1 controls delay time and R2 controls pulse width. As much as a 10:1 range can be generated.

Delay:	$C1 = 1.1 \times 10^{-5}$	T delay	c μF
Pulse:	$C2 = 1.1 \times 10^{5}$	T pulse	T μsec

PULSE GENERATOR WITH VARIABLE DUTY CYCLE



WILLIAM SHEETS

FIG. 82-7

Using only one IC and six passive components, this pulse generator has a frequency range of 400 to 4000 Hz and an adjustable duty cycle of 1 to 99%. A threshold detector (ICA) and an integrator (ICB) generate a triangular waveform. A positive voltage at the output of ICA causes the output of ICB to become a negative-going ramp. When the output of this ramp reaches a certain value, ICA, by virtue of its positive-feedback network, changes state; its output becomes negative, and the integrator generates positive ramp. This process continually repeats. A voltage follower (ICC) and a 100-kΩ potentiometer provide a variable ±0.18-V reference voltage. This reference voltage, along with the triangular waveform, feeds into the positive and negative inputs, respectively, of comparator ICD. You can set the comparator's trip voltage at any point on the triangular waveform; ICD's output changes at that point. Varying the reference voltage alters the duty cycle of the comparator's output by adjusting the potentiometer at the negative input of the integrator, thereby varying the integration time without altering the duty cycle.

83

Receiver Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE DIRECT-CONVERSION RECEIVER FOR 160 TO 20 M

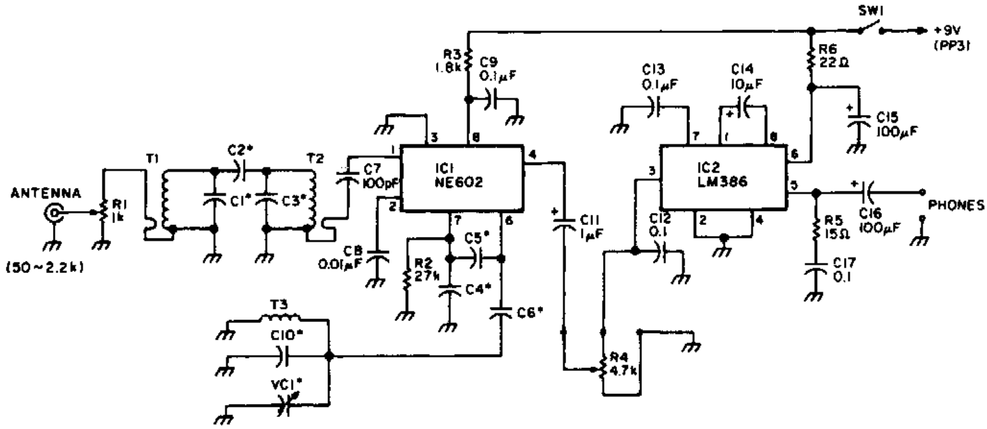


Table. Component Values for Different Bands

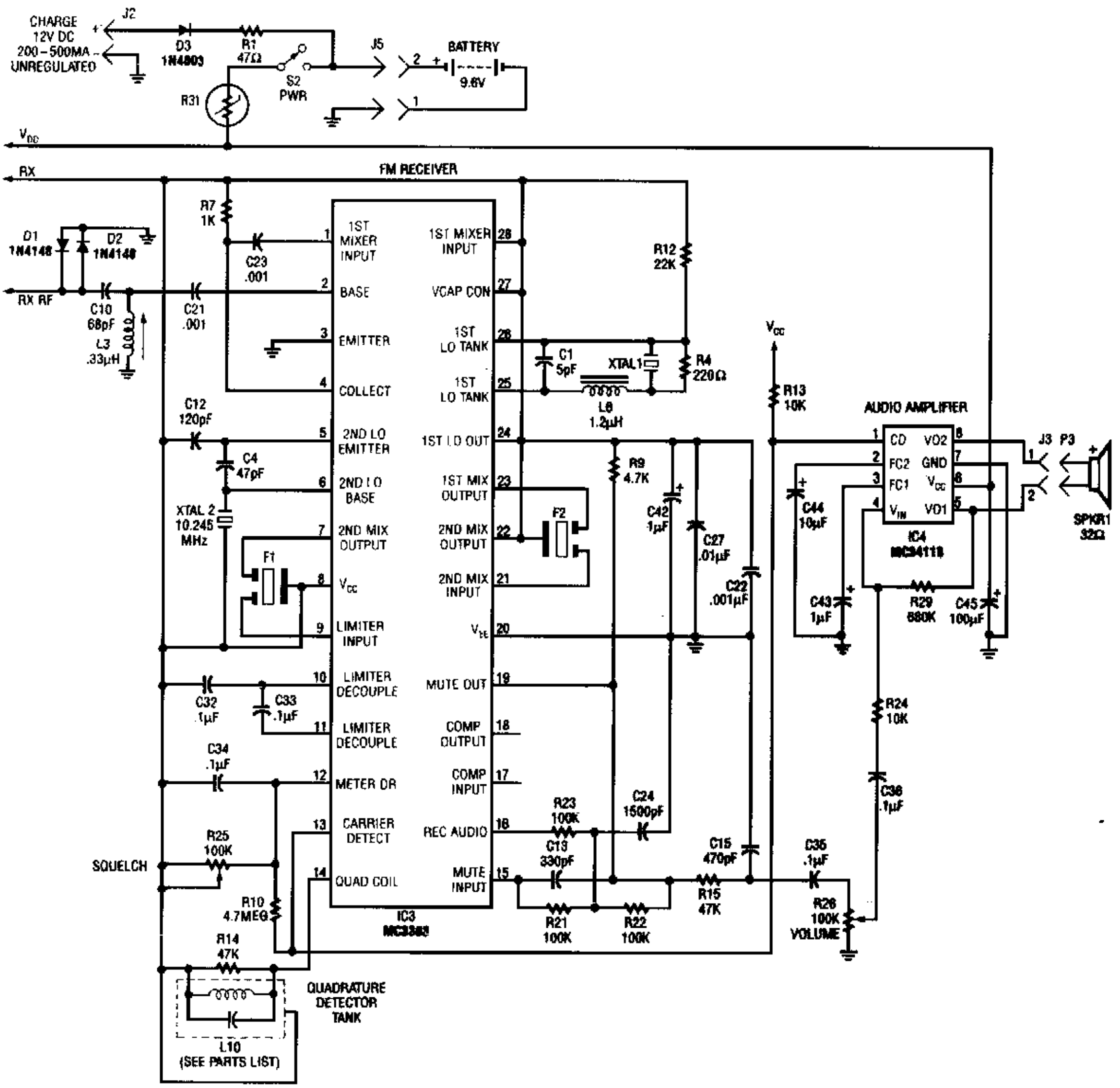
Band	C1	C2	C3	T1	T2
160	220 pF	10 pF	220 pF	BKXN-K3333R	BKXN-K3333R
80	47 pF	3 pF	47 pF	BKXN-K3333R	BKXN-K3333R
40	100 pF	8.2 pF	100 pF	BKXN-K3334R	BKXN-K3334R
30	47 pF	3 pF	47 pF	BKXN-K3334R	BKXN-K3334R
20	100 pF	3 pF	100 pF	BKXN-K3335R	BKXN-K3335R

VC1 + C10	C4	C5	C6	T3
All Sections + 100 pF	0.001 µF	0.001 µF	560 pF	BKXN-K3333R
All Sections + 100 pF	0.001 µF	0.001 µF	560 pF	BKXN-K3334R
1 Section + 47 pF	560 pF	560 pF	270 pF	BKXN-K4173AO
1 Section + 68 pF	680 pF	680 pF	220 pF	BKXN-K3335R
1 Section + 68 pF	220 pF	220 pF	68 pF	BKXN-K3335R

73 AMATEUR RADIO TODAY ***FIG. 83-1***

Note that T1 and T2 are TOKO, including part numbers for the coils T1 and T2. The direct-conversion receiver shown uses a double-tuned input network made from readily available TOKO coils. IC1, an NE602, acts as a VFO and mixer, with the output being an IF frequency in the audio range. IC2 is an audio amplifier, R4 is a volume control.

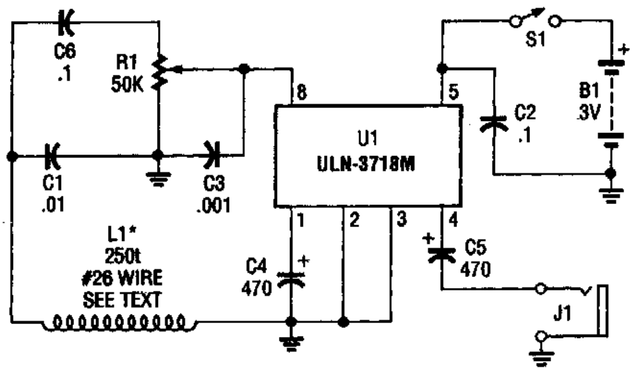
27.145-MHz NBFM RECEIVER



ELECTRONICS NOW *FIG. 83-2*

Using a Motorola MC3363 LSI one-chip FM receiver, the circuit is a dual-conversion FM receiver with a 10.7-MHz IF chain. IC4 provides power to drive a small speaker.

VLF WHISTLER RECEIVER

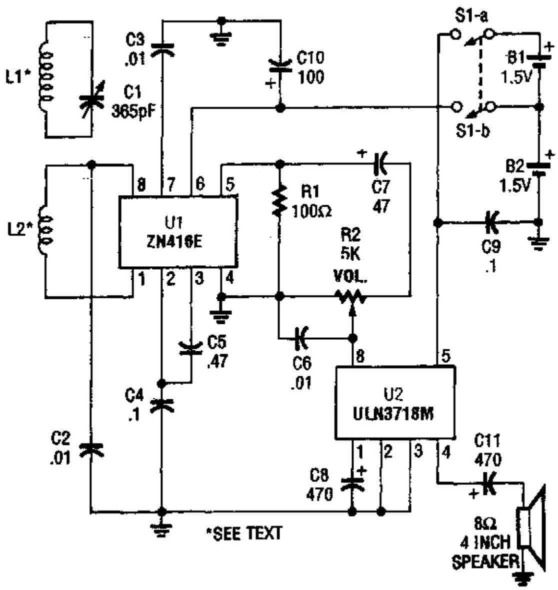


POPULAR ELECTRONICS

FIG. 83-3

The VLF whistler receiver is intended to listen to natural radio noise and signals that occur below 20 kHz. L1 is a large loop antenna that is 250 to 300 turns #26 gauge wire on a form 3' diameter. L1 should be mounted well away from power lines and is oriented for minimum 60- and 120-Hz pickup.

BASIC AM RECEIVER CIRCUIT

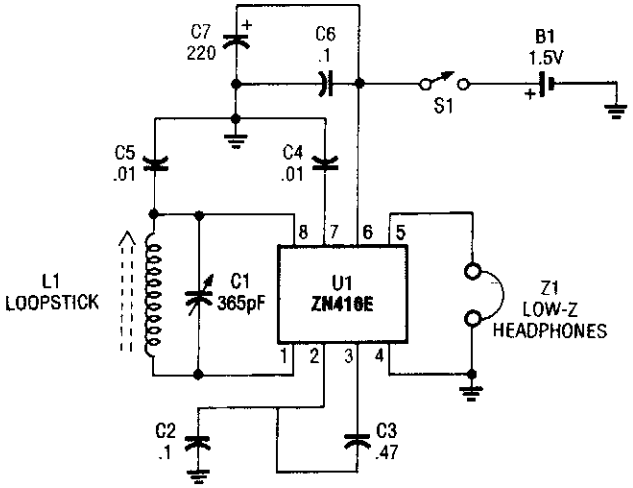


POPULAR ELECTRONICS

FIG. 83-4

Using a single ZN416E IC and a ULN3718M, this simple TRF receiver can drive a loudspeaker. Two 1.5-V cells power the circuit.

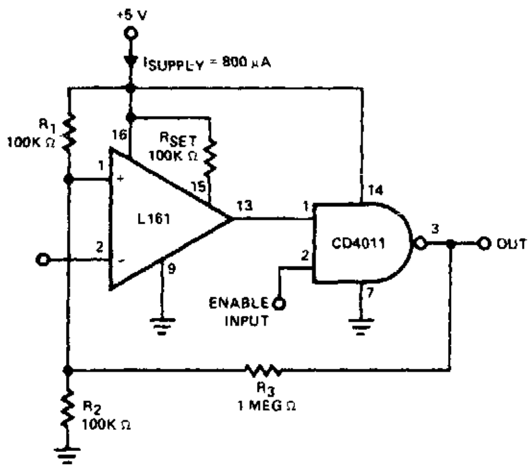
SIMPLE 1.5-V AM BROADCAST RECEIVER



POPULAR ELECTRONICS ***FIG. 83-5***

This receiver uses the ZN416E made by GEC Plessey. The tuning is via C1.

CMOS LINE RECEIVER



INTEGRATED CIRCUITS DATA BOOK ***FIG. 83-6***

This circuit will interface a line input to CMOS. The supply current is >1 mA at +5 V.

NE602 DIRECT-CONVERSION RECEIVER

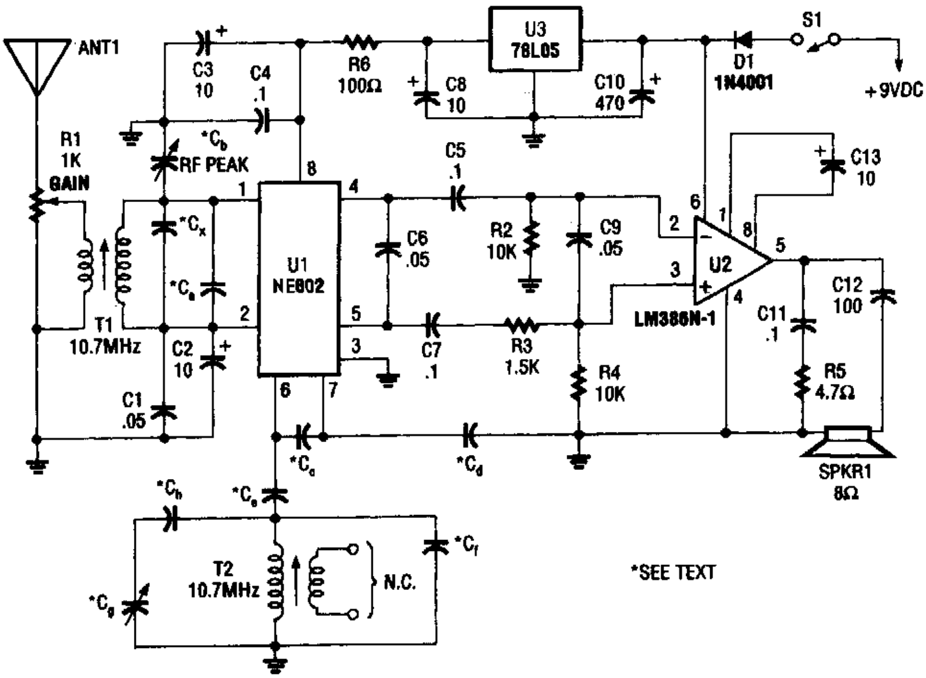


Table 1-- CAPACITOR SELECTION

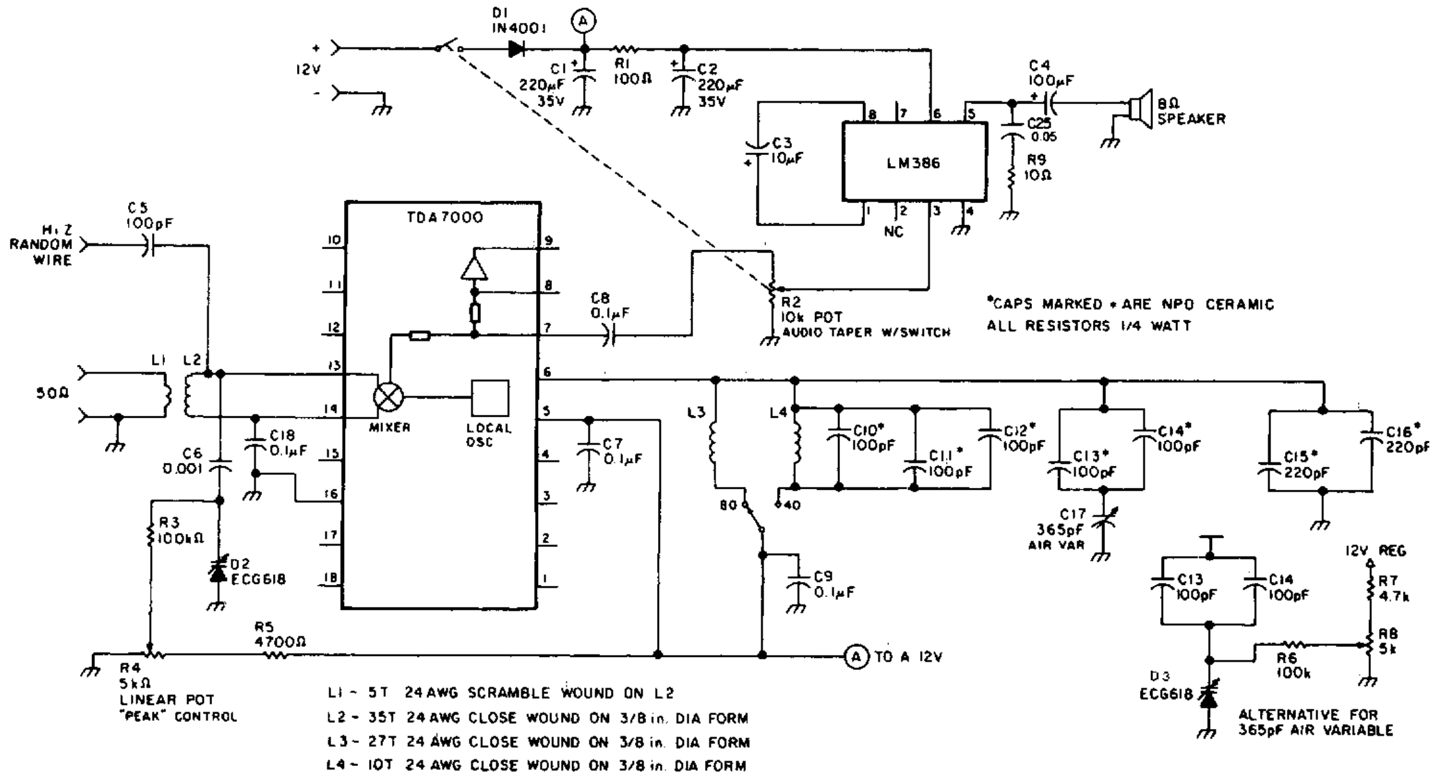
Band (meters)	Capacitor values (picofarads)					
	Cc	Cd	Ce	Cf	Cg	Ch
75/80	1000	1000	470	120	365	270
40	330	330	120	150	365	68

POPULAR ELECTRONICS

FIG. 83-7

An NEC602 is used as a mixer with a zero IF frequency output. U2 acts as an audio amplifier. This receiver is primarily for SSB and CW signals. T1 and T2 are 10.7-MHz IF coils used in AM/FM transistorized radios, etc. or in any similar indicator.

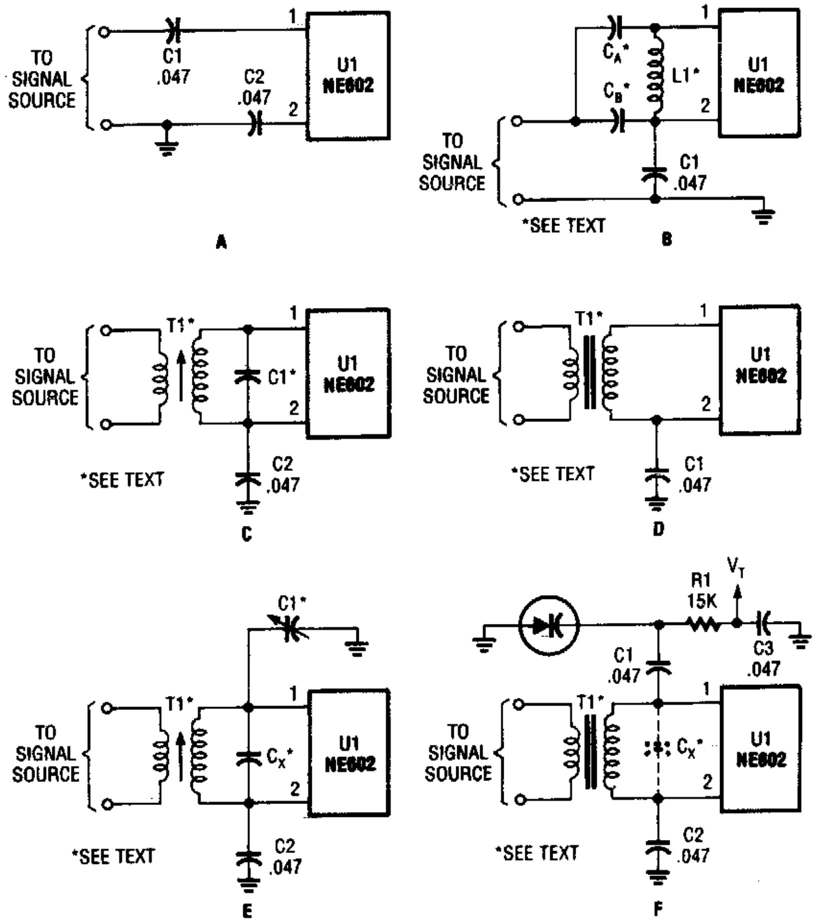
80- AND 40-M CW/SSB RECEIVER



73 AMATEUR RADIO TODAY ***FIG. 83-8***

This direct-conversion receiver uses a TDA7000 IC and it drives an LM386 audio amplifier. The TDA7000 is used for its mixer and L.O. section. The frequency control can be either with an air variable capacitor or a varactor diode.

NE602 RF INPUT CIRCUITS

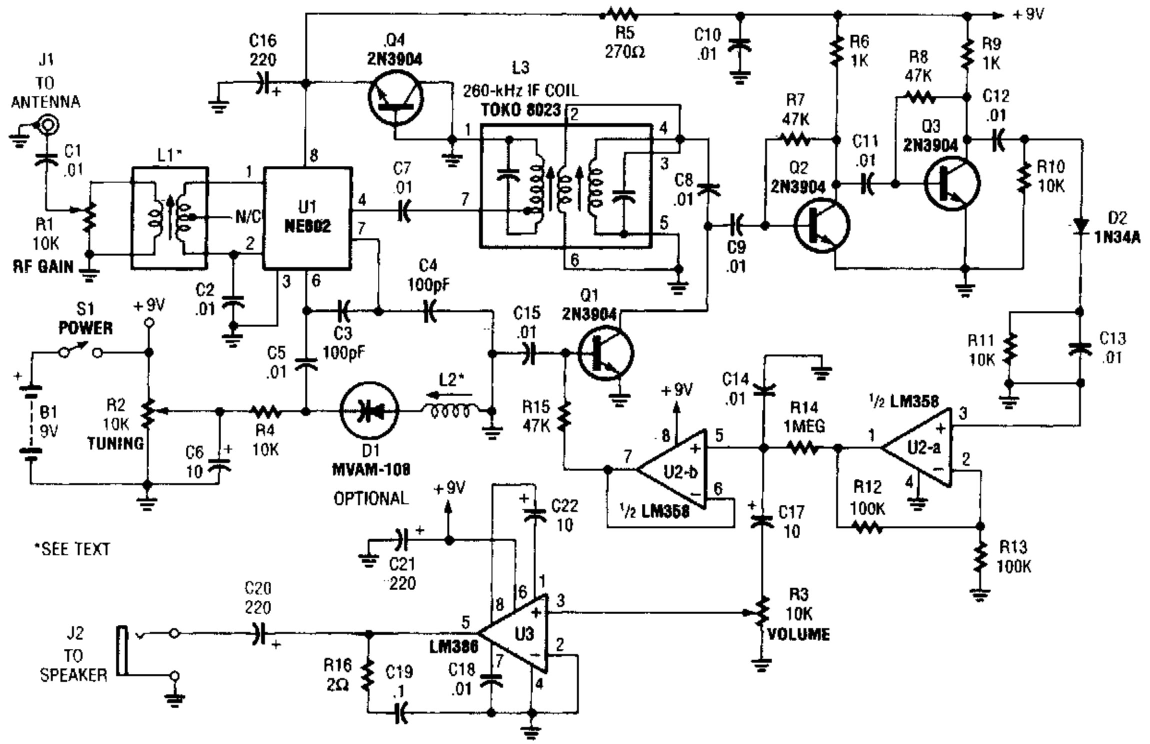


POPULAR ELECTRONICS

FIG. 83-9

Here are a few of the many possible RF input circuits for the NE602. Just about any tuned or broadband circuit will work.

SUPER-SIMPLE SHORTWAVE RECEIVER



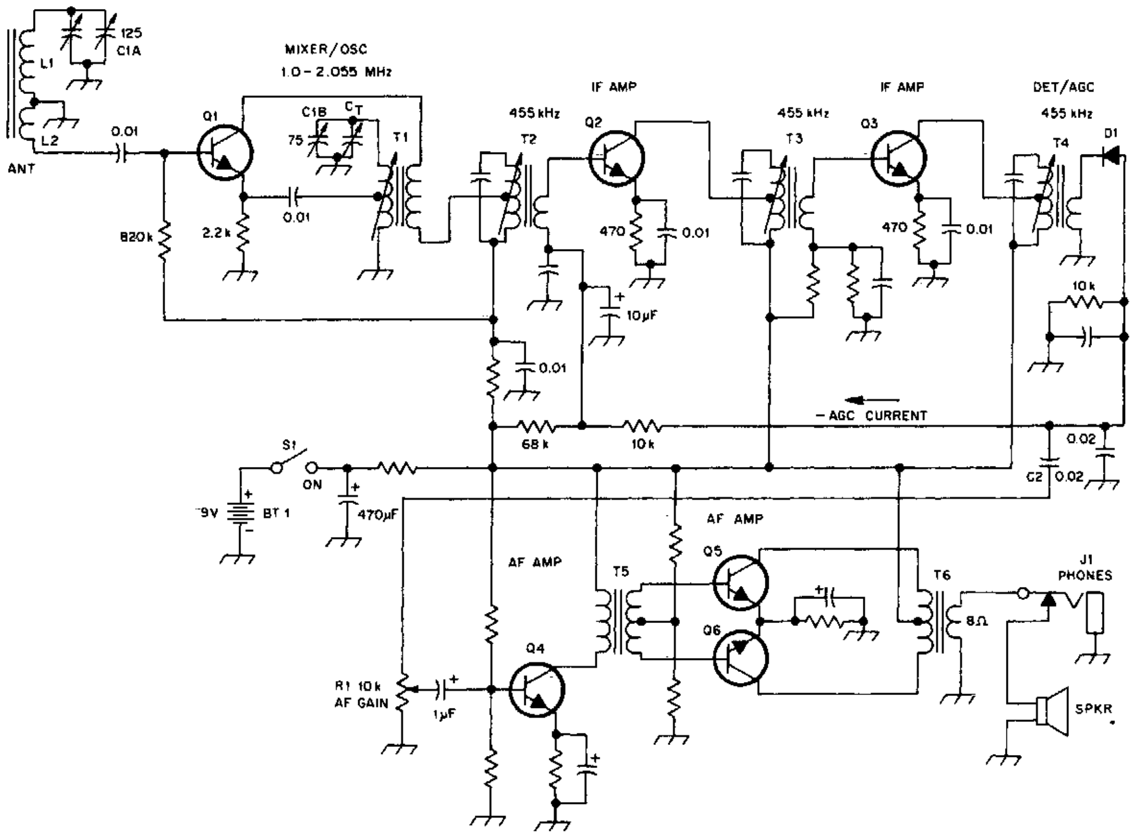
POPULAR ELECTRONICS

FIG. 83-10

Integrated circuit U1 (an NE602 double-balanced mixer) is a combination oscillator and frequency mixer. Signals from the antenna input (at J1) are fed through dc-blocking capacitor C1 to the RF-gain control, R1, and fed to the input of U1 at pins 1 and 2.

The local-oscillator frequency, which varies with the settings of R2 and L2, is mixed internally within U1, resulting in an output. The mixer output at pin 4 of U1 is applied to a tunable 260-kHz band-pass intermediate-frequency (IF) transformer, L3, through dc-blocking capacitor C7. Therefore, signals that are roughly 260 kHz above and below the local-oscillator frequency are passed while others are effectively blocked. The IF frequencies are now amplified by Q2 and Q3. The AM audio signal is detected by D2 and its associated components, which bypass the RF signals, and leave only the audio signals. The signals are preamplified by U1-a (half of an LM358 dual op amp). The audio is then boosted to speaker level by the LM386 low-voltage audio power amplifier, U3.

TRANSISTORIZED AM RADIO

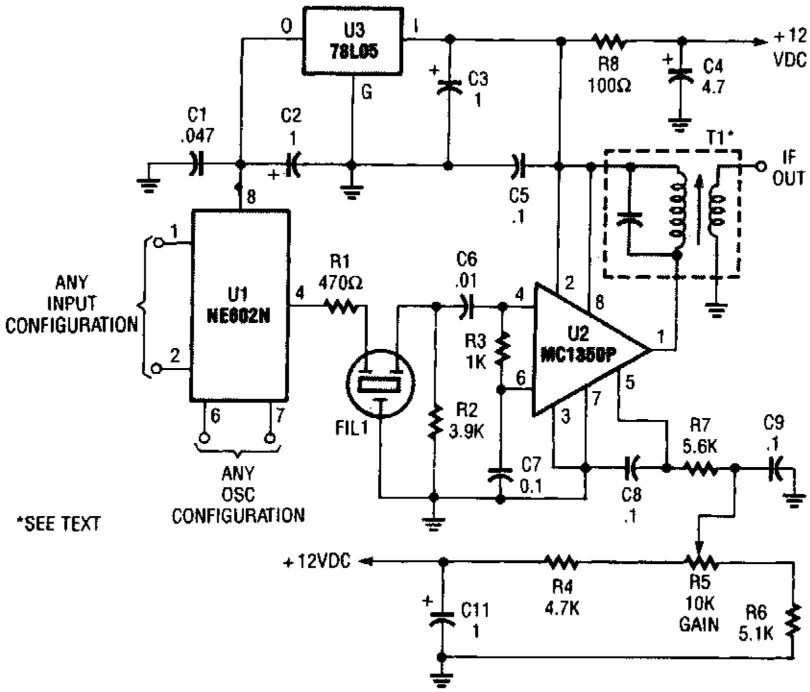


QST

FIG. 83-11

Shown is a schematic of a typical transistor AM radio. This circuit uses npn transistors. The circuit is "generic;" therefore, no specific values are given for some components. This circuit is for reference, to serve as a starting point for experimenters.

NE602 SUPERHET FRONT END



POPULAR ELECTRONICS

FIG. 83-12

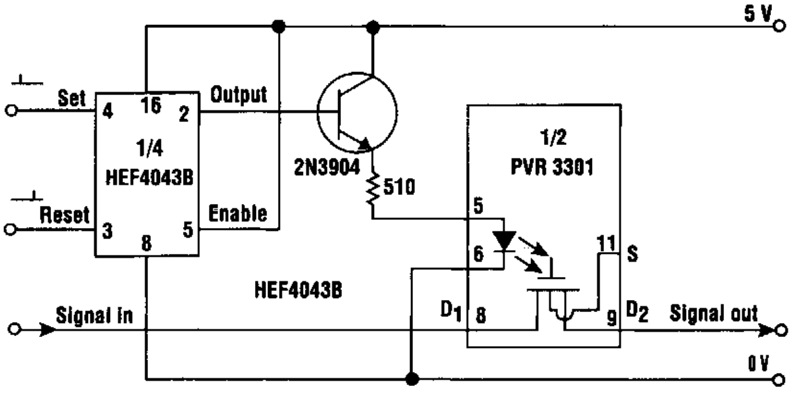
By using an NE602 with a filter and an MC1350P IC, a front end and an IF system for a basic superheterodyne receiver can be built with few parts. T1 is any suitable IF transformer for 262 kHz, 455 kHz, 10.7 MHz, etc.

84

Relay Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SOLID-STATE LATCHING RELAY

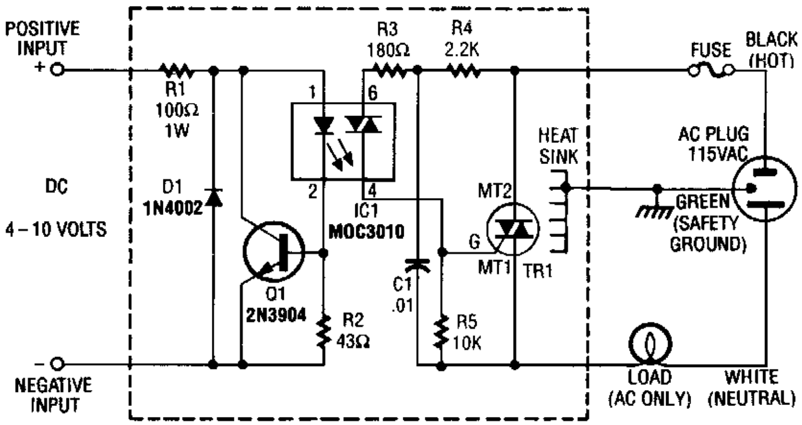


ELECTRONIC DESIGN *FIG. 84-1*

This simple circuit provides a solid-state equivalent of the electromechanical latching relay (see the figure). What's more, the switching is clean, highly resistant to vibration and shock, and isn't sensitive to magnetic fields or position.

The circuit operates as follows: a set pulse to the 4043 RS latch takes its output high and turn on the 2N3904 transistor. Current will then flow through the photovoltaic relay's LED and the resistance between D1 and D2 will fall from several gigaohms to less than 30 Ω. The PVR will remain in this state until a reset pulse is received by the 4043 RS latch.

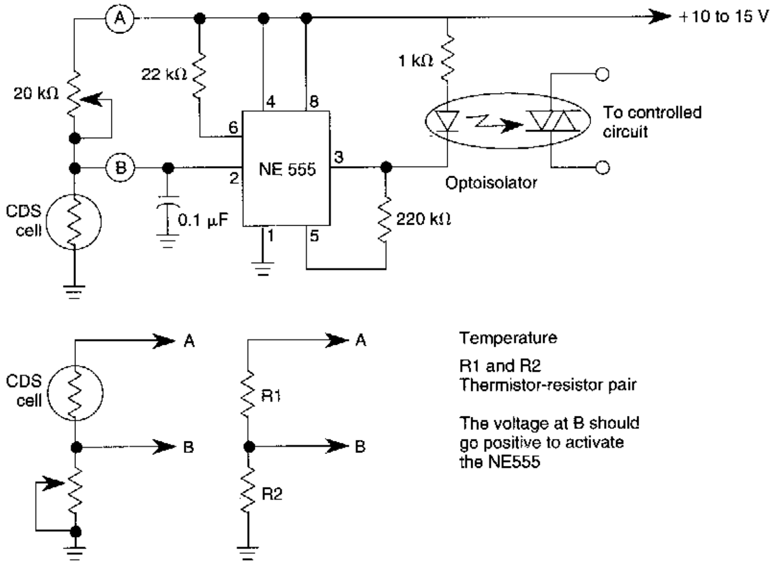
SOLID-STATE RELAY CIRCUIT



RADIO ELECTRONICS *FIG. 84-2*

R1 limits input current while Q1 acts as a current sink to protect IC1. D1 serves as a polarity protector. IC1 provides a triac output to trigger the main triac, TR1.

SOLID-STATE RELAY CIRCUITS

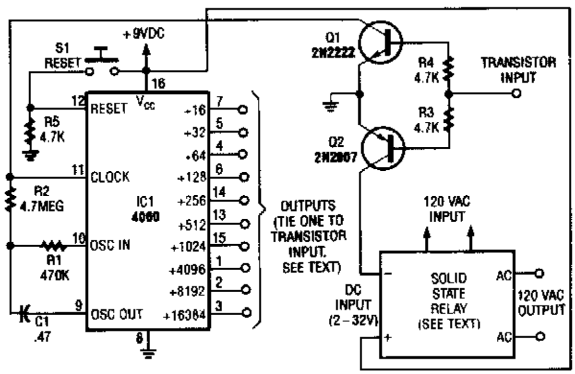


WILLIAM SHEETS

FIG. 84-3

This dark-activated relay switch can be used to turn on walkway or other outdoor lighting at dusk. By using alternate connections to A and B, increasing illumination, high and low temperatures can be sensed.

TIME DELAY RELAY

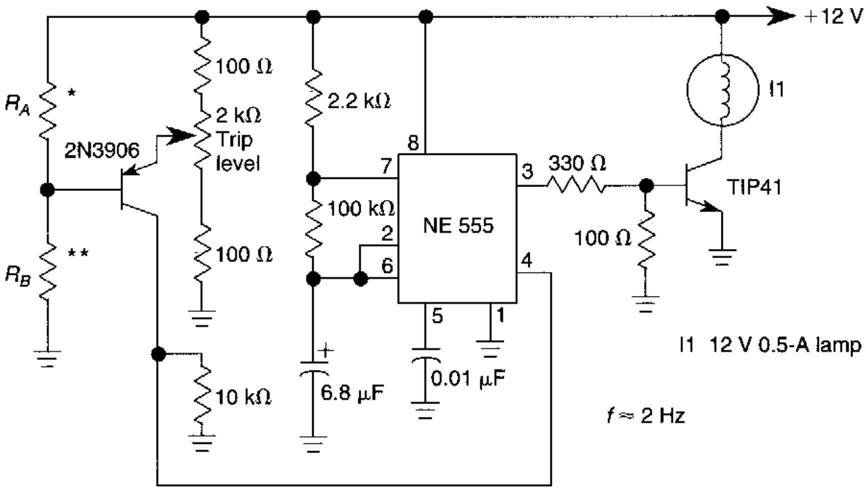


ELECTRONICS NOW

FIG. 84-4

Using a 4060 CMOS binary divider and built-in clock oscillator, a long-duration timer can be made very simply. The solid-state relay can be sized for your application, and can be replaced with a mechanical relay if a suitable power supply is available. With the components shown, a 4.5-Hz clock frequency is generated. Divided outputs are available from ÷ 4 to 16384 (about 4 hours).

SENSOR-ACTIVATED RELAY PULSER



Either R_A or R_B can be sensors, as desired. A decrease in R_B or an increase in R_A will cause the NE555 to flash I1. R_A and R_B should be ≤100 kΩ max.

WILLIAM SHEETS ***FIG. 84-5***

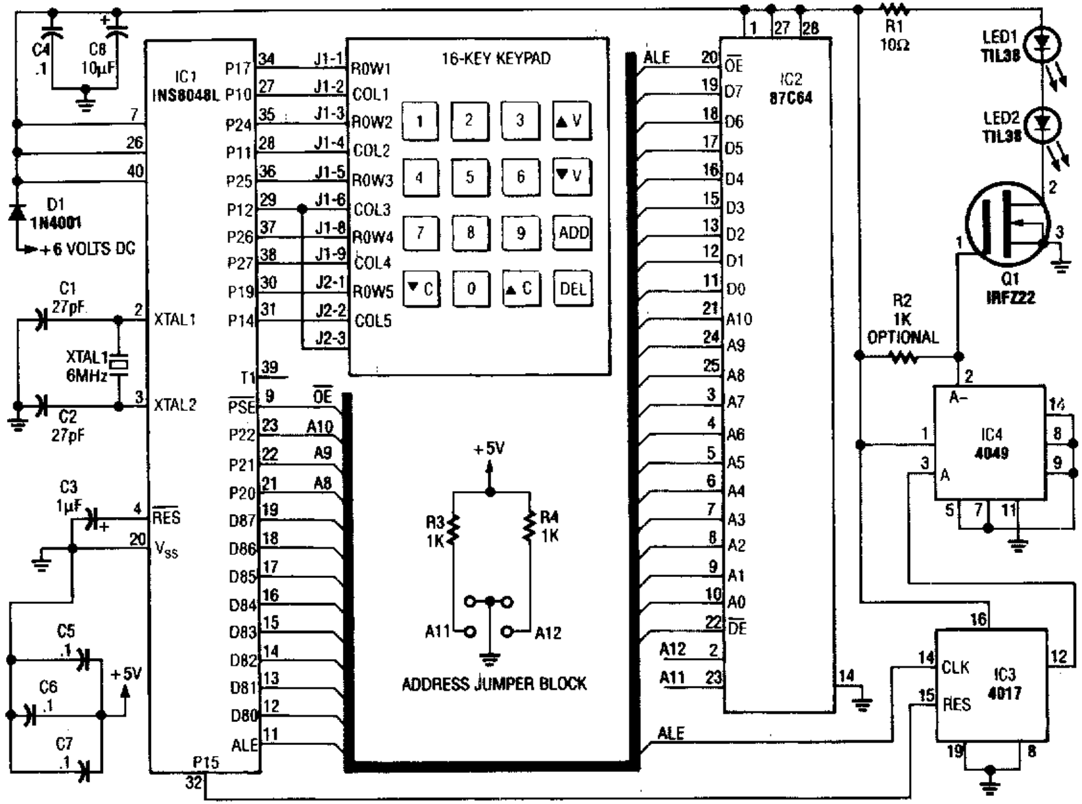
A sensor turns on Q1 to activate the low-frequency 555 oscillator, which pulses LAMP I1. Sensor may be sensitive to changes in light or temperature.

85

Remote-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

REMOTE-CONTROL TRANSMITTER



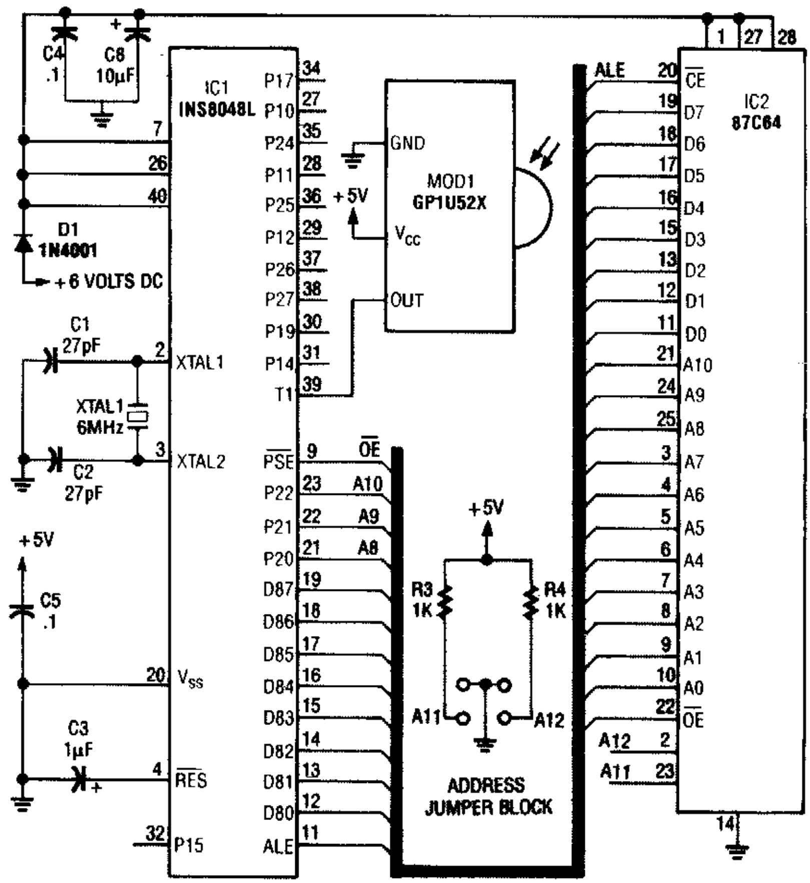
IR TRANSMITTER SCHEMATIC. The 40-kHz carrier is derived by dividing IC1's oscillator frequency (6 MHz) by 15, to get 400 kHz, which is divided by 10 by IC3.

ELECTRONICS NOW

FIG. 85-1

This transmitter sends an FM signal in the 88-to 108-MHz range, with a tone of 19 kHz. This can be used to activate the FM MPX pilot carrier indicator, which can be interfaced to external devices. L4 is for use with a 15 CM wire antenna. L1 is 9 turns of #26 enamelled wire on a ¼-W 10-kΩ resistor (carbon type), L2 is 2 turns wound over L1. L3 is 7 turns of #26 enamelled wire on a 10-kΩ ¼-W resistor.

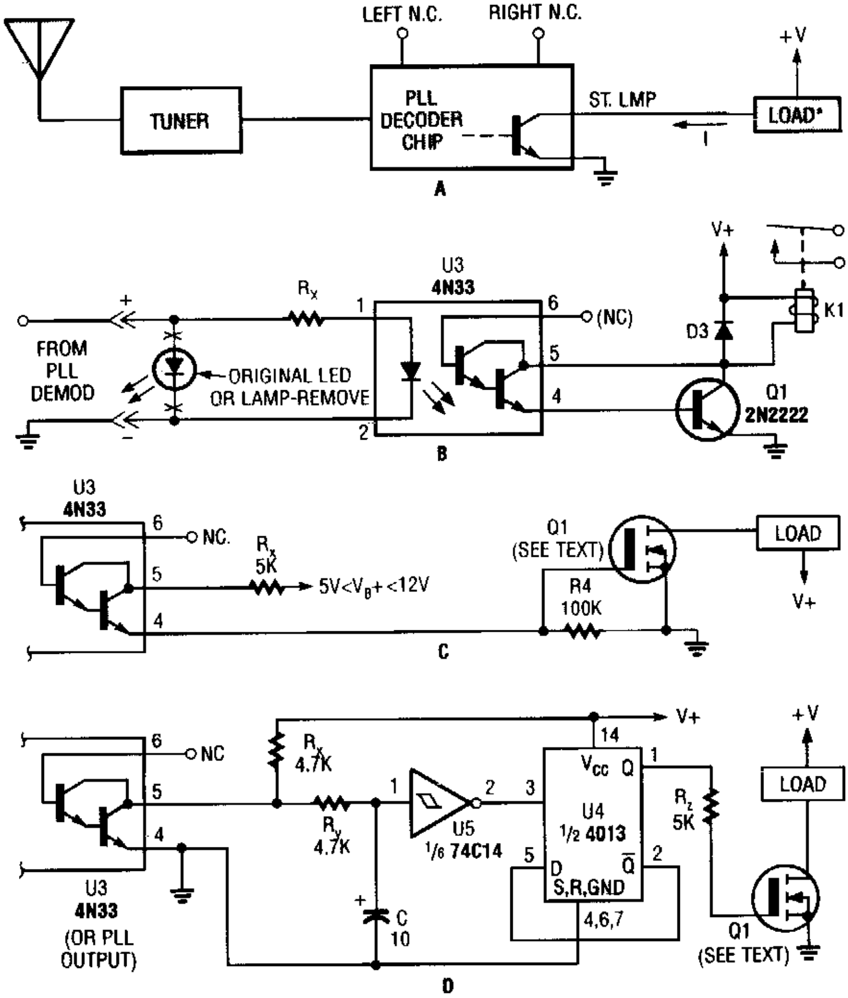
REMOTE-CONTROL RECEIVER



ELECTRONICS NOW *FIG. 85-2*

This circuit is based on the Sharp GP1U52X IR module and INS8048L microprocessor. The GP1U52X is a hybrid IC/infrared detector that provides a strong clean signal for later filtering and demodulation.

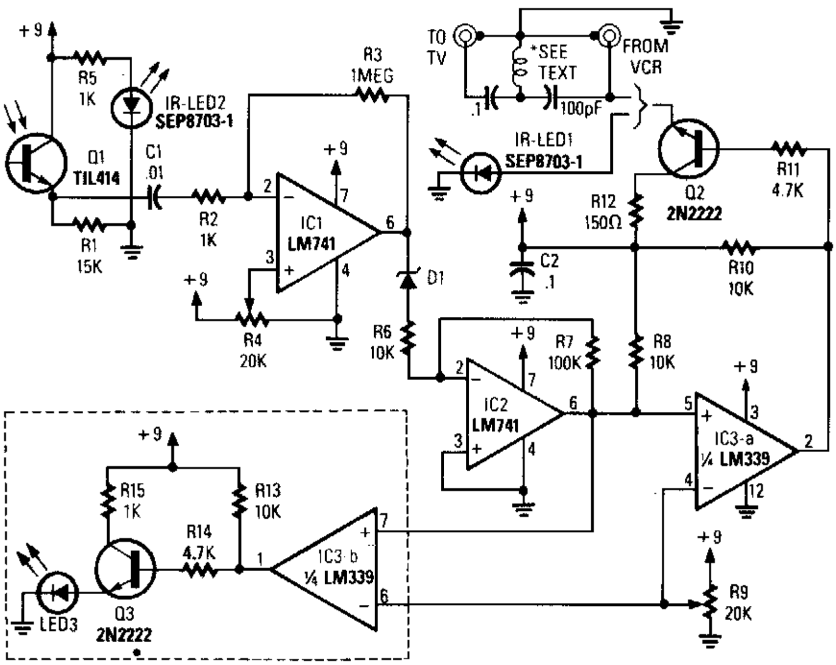
INTERFACE CIRCUITS FOR THE REMOTE-CONTROL TRANSMITTER



POPULAR ELECTRONICS *FIG. 85-3*

Shown here are several possible interface circuits that can be used with the remote-control transmitter. The one in A illustrates a typical FM stereo MUX decoder with a load connected directly to the open-collector output of a TA7343 PLL. The circuit in B illustrates an optoisolator-coupler output driving a 12-V relay coil via a general-purpose transistor. C shows the gate of an N-channel power MOSFET connected to the output of a 4N33. The final circuit, D, is a toggle flip-flop that allows push-on/push-off control.

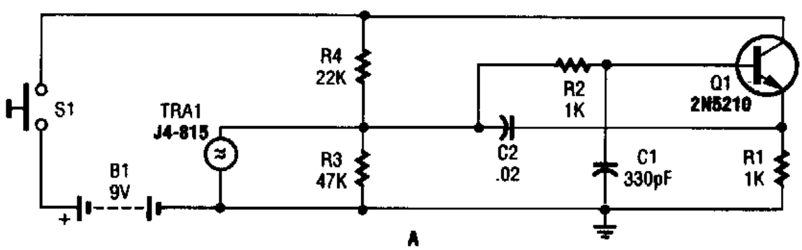
REMOTE-CONTROL EXTENDER



1991 R-E EXPERIMENTERS HANDBOOK ***FIG. 85-4***

A signal from an IR remote control is converted from IR radiation to a frequency pulse that can be transmitted through coaxial TV cable or any other two-conductor wire to another room, where it's converted back into an IR signal.

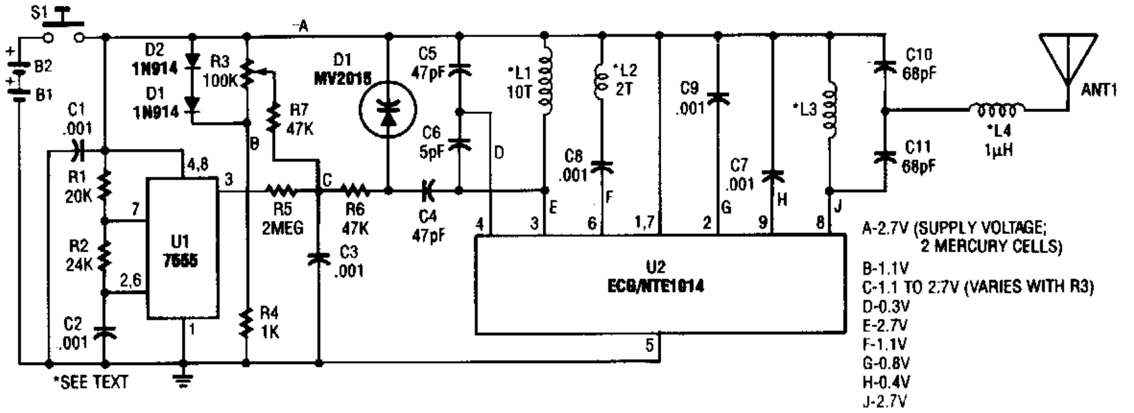
ULTRASONIC REMOTE-CONTROL TRANSMITTER



POPULAR ELECTRONICS ***FIG. 85-5***

A GC Electronic P/N J4-815 ultrasonic transducer is used in this 40-kHz transmitter for remote-control application.

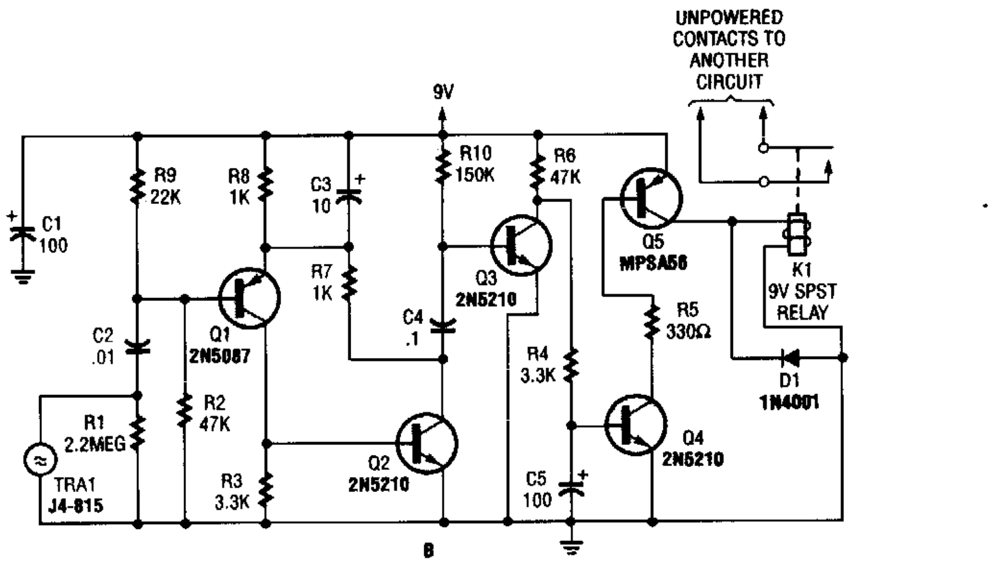
REMOTE-CONTROL TRANSMITTER



POPULAR ELECTRONICS *FIG. 85-6*

This transmitter can be used for a variety of purposes. An INS8048L microprocessor generates various codes depending on keypad presses. The codes are modulated on a 40-kHz carrier. Q1 drives IR LEDs LED1 and LED2.

ULTRASONIC REMOTE-CONTROL RECEIVER



POPULAR ELECTRONICS *FIG. 85-7*

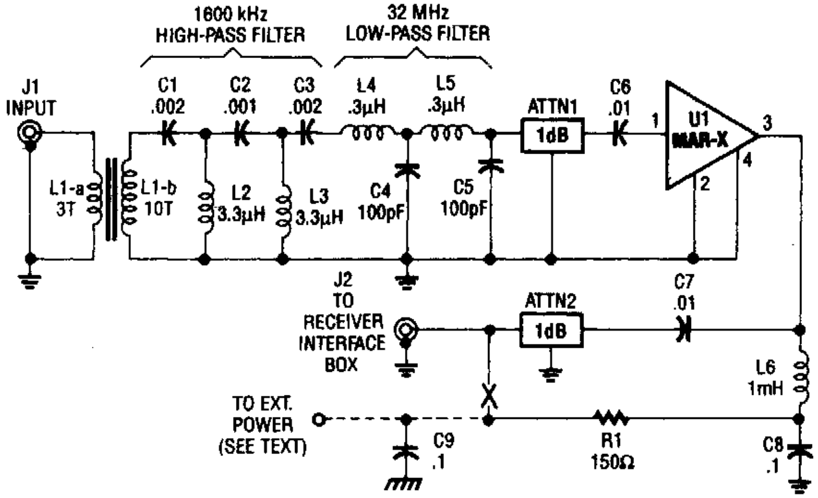
A GC Electronics P/N J4-815 transducer is used to receive 40-kHz acoustic remote-control signals. The receiver drives a relay for control of another circuit.

86

RF Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

HF PREAMPLIFIER

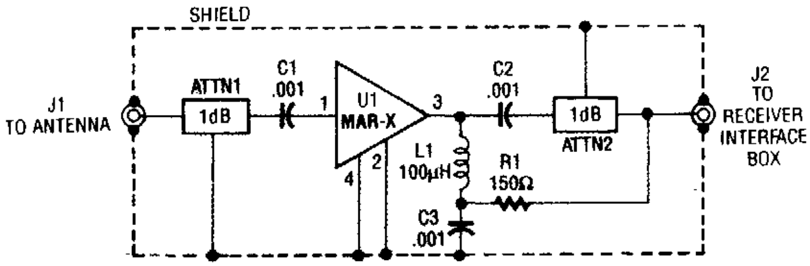


POPULAR ELECTRONICS

FIG. 86-1

This HF SW receiver preamplifier is comprised of a broadband toroidal transformer (L1-a and L1-b), a complex LC network (comprised of a 1600-kHz, high-pass filter and a 32-MHz, low-pass filter), L2 and L3 (26 turns of #26 enameled wire wound on an Amidon Associates T-50-2, red, toroidal core), a pair of resistive attenuators (ATTN1 and ATTN2), and of course, the MAR-x device. External power for the preamp can be 9 to 12 Vdc. R1 can be increased in value for higher voltages.

VHF/UHF PREAMP USING MAR-x

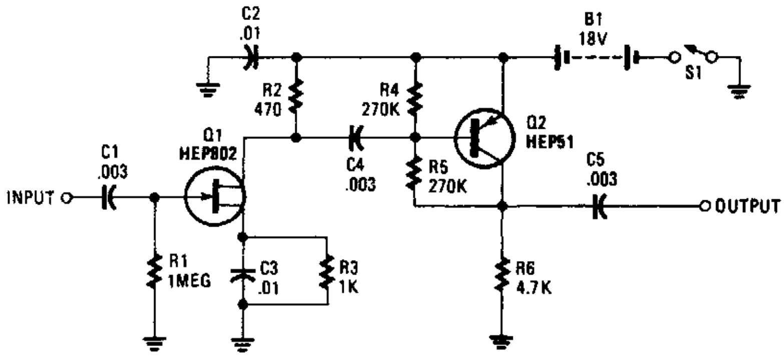


POPULAR ELECTRONICS

FIG. 86-2

The MAR-x preamp shown will cover up to 1.5 or 2 GHz with the correct MAR-x IC. ATTN1 should be omitted for low noise-figure applications. ATTN1 and ATTN2 provide a means of limiting possible termination range, for less chance of device instability.

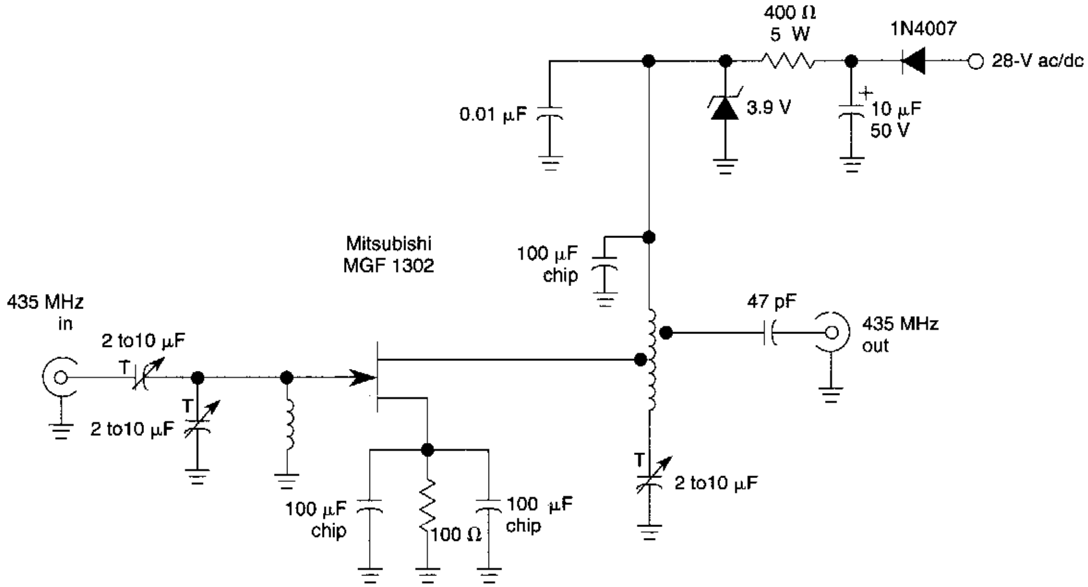
BROADBAND RF AMPLIFIER



POPULAR ELECTRONICS **FIG. 86-3**

The use of a FET gives this amplifier a high input impedance. The bandwidth should be adequate for LW through HF use (dc-30 MHz), as an active antenna preamplifier.

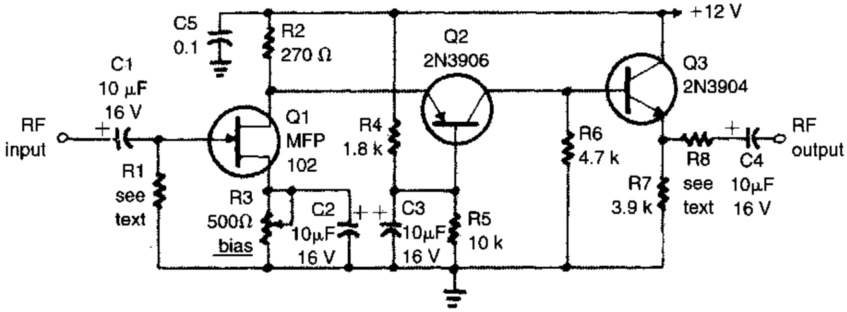
LOW-NOISE GASFET PREAMP FOR 435 MHz



WILLIAM SHEETS **FIG. 86-4**

This circuit is a low-noise preamplifier for the 435-MHz amateur satellite frequencies. The circuit uses a Mitsubishi MGF1302. A 28-Vdc source is shown, although by changing the 400-Ω 5-W resistor lower voltages can be used.

BROADCAST-BAND RF AMPLIFIER

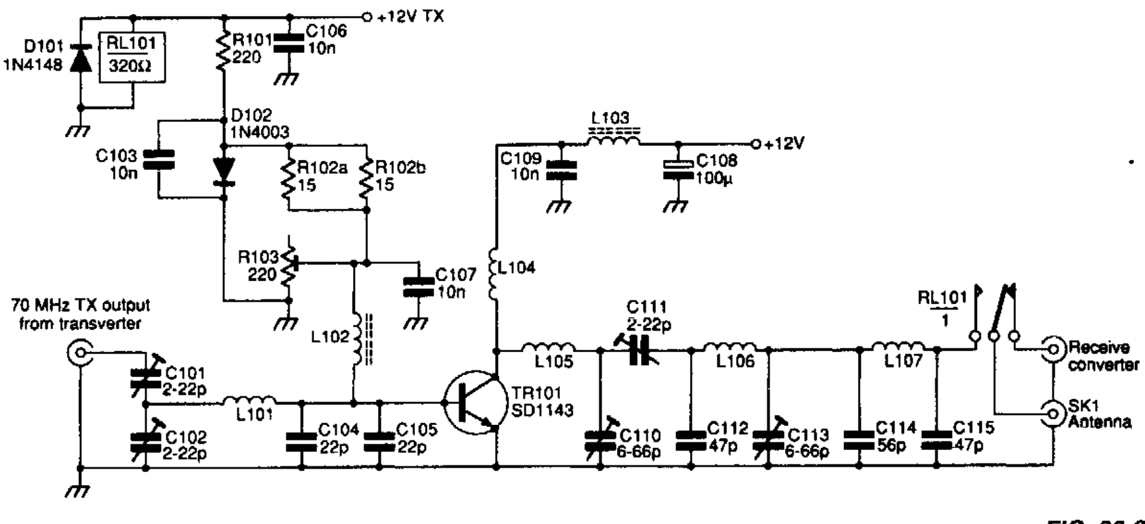


R-E EXPERIMENTERS HANDBOOK

FIG. 86-5

The circuit has a frequency response that ranges from 100 Hz to 3 MHz; the gain is about 30 dB. Field-effect transistor Q1 is configured in the common-source self-biased mode; optional resistor R1 allows you to set the input impedance to any desired value. Commonly, it will be 50 Ω. The signal is then direct-coupled to Q2, a common-base circuit that isolates the input and output stages and provides the amplifier's exceptional stability. Last, Q3 functions as an emitter-follower, to provide low output impedance (about 50 Ω). If you need higher output impedance, include resistor R8. It will affect impedance according to this formula: $R_8 \approx R_{OUT} - 50$. Otherwise, connect output capacitor C4 directly to the emitter of Q3.

70-MHz RF POWER AMPLIFIER

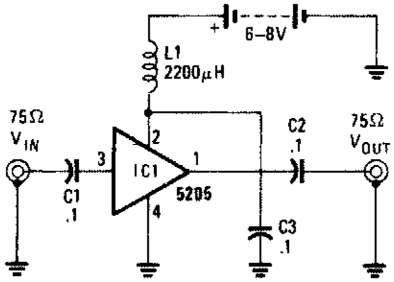


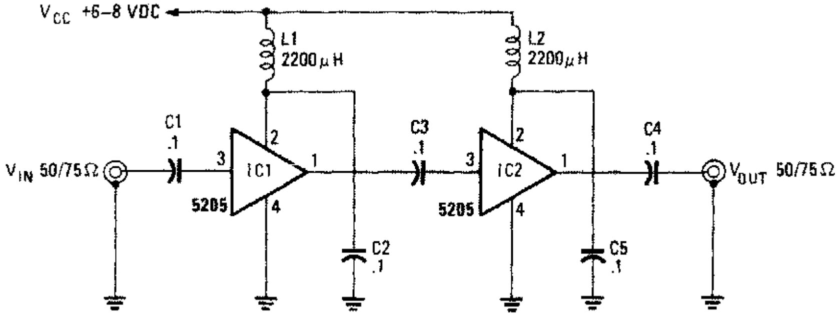
PRACTICAL WIRELESS

FIG. 86-6

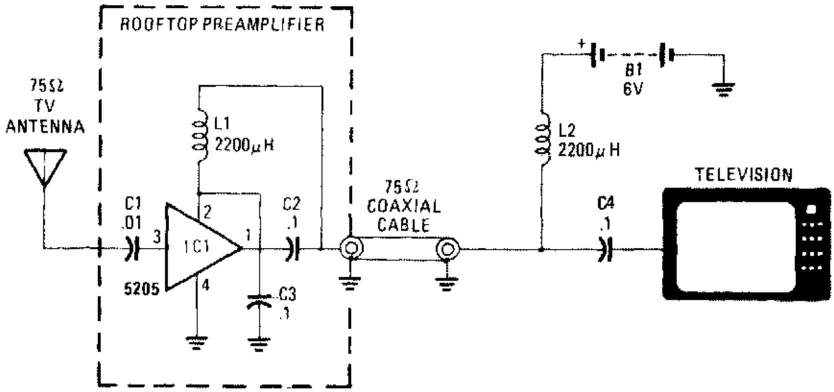
The SD1143 transistor provides a gain of about 14 dB in this circuit. It uses the fact that a 175-MHz device has a much higher gain when used at lower frequencies. The amplifier was originally designed to be used with a transverter. The output is 8 to 10 W for a 300- to 500-mW input.

MINIATURE WIDEBAND AMPLIFIER





SINCE THE NE5205 FUNCTIONS as a gain block, two or more can be easily cascaded to provide additional amplification. In this circuit, which uses two NE5205s, the overall gain is 40 dB.

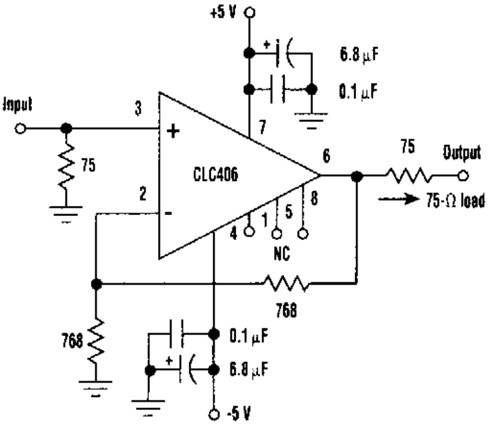


IF THE POWER SUPPLY is fed through the signal-carrying coaxial cable, the amplifier can be mounted in a weatherproof enclosure directly at the antenna.

R-E EXPERIMENTERS HANDBOOK ***FIG. 86-7***

Except for the coupling and decoupling capacitors, IC1 is a complete wideband amplifier that has a fixed gain of 20 dB to 450 MHz. No external compensation is required.

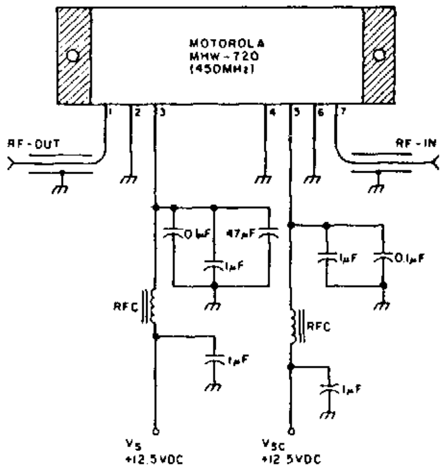
30-MHz AMPLIFIER



ELECTRONIC DESIGN *FIG. 86-8*

Using a CLC406 op amp, this video amplifier has a voltage gain of +2 and is flat to 30 MHz. The circuit should be useable in video switching and interfacing applications.

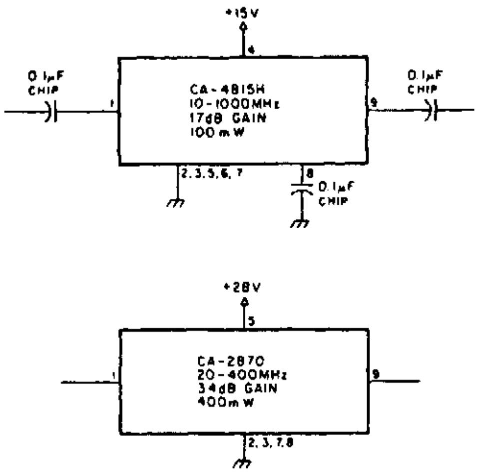
20-W 450-MHz AMPLIFIER



73 AMATEUR RADIO *FIG. 86-9*

Delivering 20-W output, this amplifier has a gain of 21 dB at 450 MHz. A 12-V supply powers this circuit.

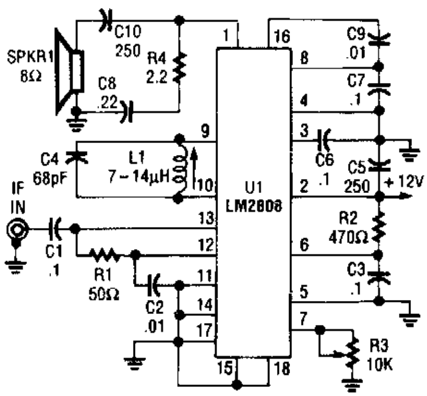
WIDEBAND POWER AMPLIFIER



73 AMATEUR RADIO *FIG. 86-10*

Using TRW P/N CA-815H, a 17-dB gain amplifier that delivers 100 mW over 10 to 1000 MHz can be constructed. The CA-2870 will yield 0.4 W with 34-dB gain from 20 to 400 MHz.

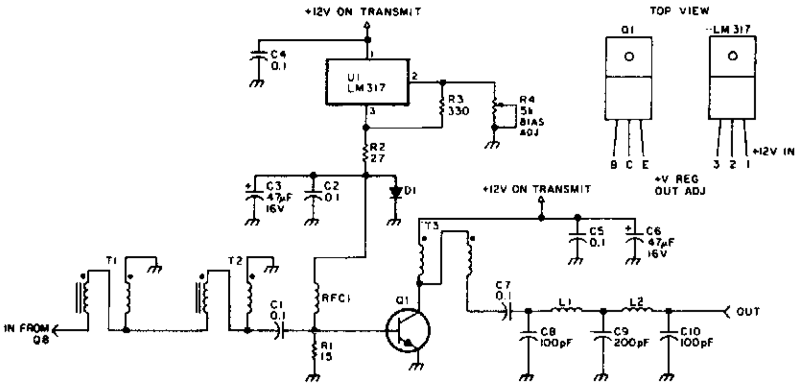
TV SOUND SYSTEM

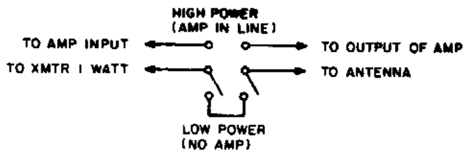


POPULAR ELECTRONICS *FIG. 86-11*

An LM2808 performs IF amplification of the 4.5-MHz sound subcarrier, limiting, detection, and audio amplification. If the center frequency must be changed, then change L1/C4. Audio output is 0.5 W. R3 is the volume control.

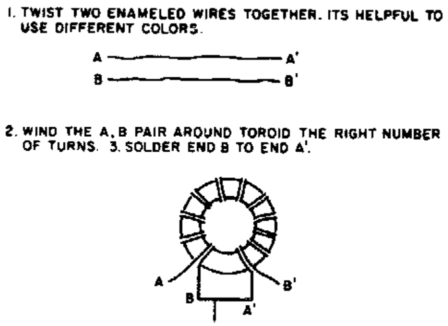
10-W 10-METER LINEAR AMPLIFIER





HI/LO POWER SWITCH

A double-pole double-throw switch can be used to switch the amplifier in and out of the circuit.



Bifilar winding details for T1, T2 and T3.

Table 1. Output filter values for other bands.

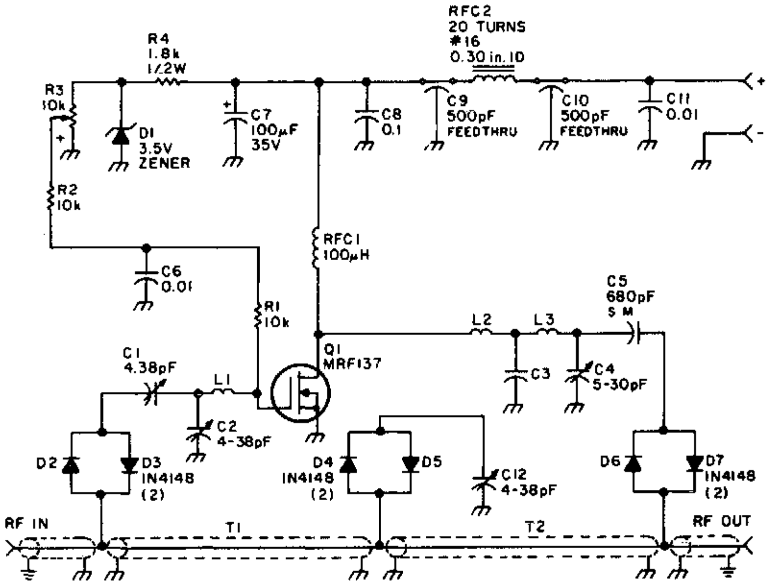
Band (meters)	C1,C3	C2	L1,L2
12	117 pF	220 pF	8 turns, T-50-6 toroid
15	138 pF	270 pF	9 turns, T-50-6 toroid
20	138 pF	420 pF	12 turns, T-50-6 toroid
30	289 pF	579 pF	12 turns, T-50-2 toroid
40	400 pF	800 pF	14 turns, T-50-2 toroid
80	700 pF	1415 pF	19 turns, T-50-2 toroid

Note: use #26 wire for C1 and C2. Use capacitors that are closest to these suggested values. As the operating frequency decreases, the gain will increase as well as the possibility for instability. You may have to use RC feedback to negate this effect. Values for the above table were obtained from the QRP Notebook by Doug DeMaw.

73 AMATEUR RADIO TODAY **FIG. 86-12**

This linear amplifier delivers 10-W PEP output with 1.25-W drive on 10 m. T1, T2, and T3 are 10 turns of bifilar windings on an FT-50-43 toroidal core. The transformers are broadband. Filters for other bands, if desired, are shown.

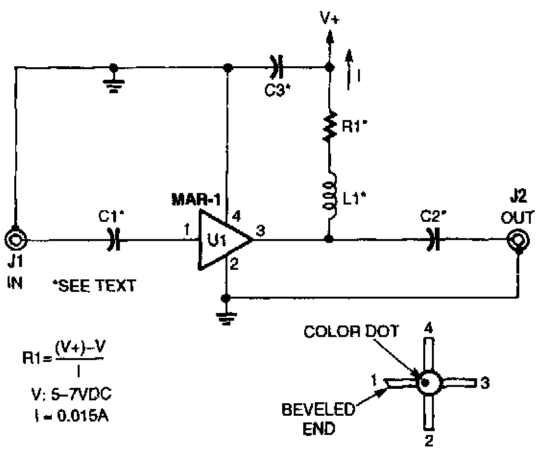
2-METER FET POWER AMPLIFIER FOR HTs



73 AMATEUR RADIO TODAY ***FIG. 86-13***

Using a power MOSFET, this amplifier can boast a 2-W handie-talkie power level to around 10 W on 2 meters. A transmission-line RF switch is used for T/R switching.

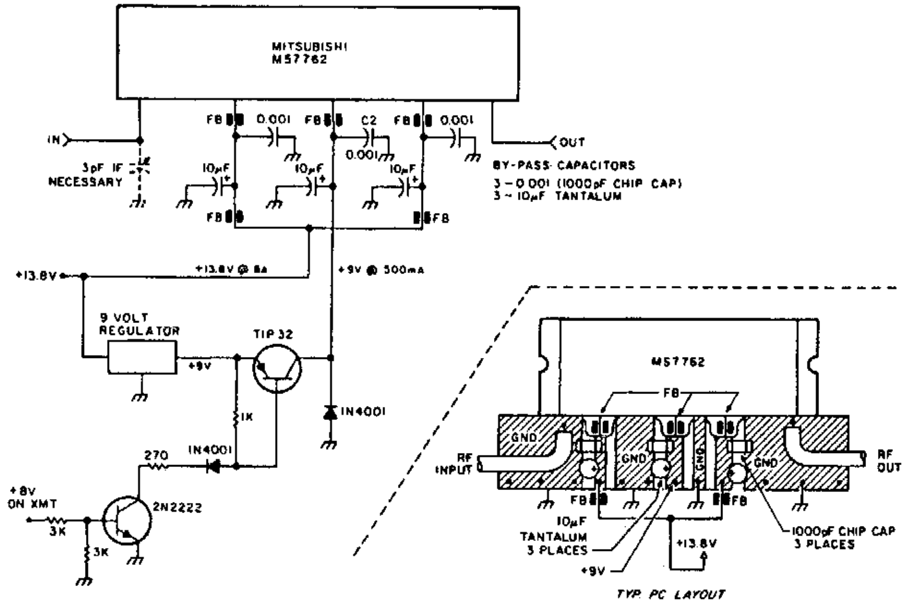
RECEIVER/SCANNER PREAMP USING MAR-1 MMIC



The low-cost Mini-Circuits MAR-X series of chips offer the RF builder a real advantage, with their inherent 50-Ω input and output impedances (needed for RF systems). An MAR-1-based receiver/scanner preamplifier is shown. C1 and C2 are chip capacitors. Use 0.01 µF for HF, 0.001 for VHF, and 100 pF for above 100 MHz, depending on the low-frequency limit that you desire. C3 can be a ceramic disc of 0.01 µF or 0.001 µF, depending on frequency range. L1 is an RF choke that is suitable for the frequency range that you desire (0.1 to 10 µH).

POPULAR ELECTRONICS ***FIG. 86-14***

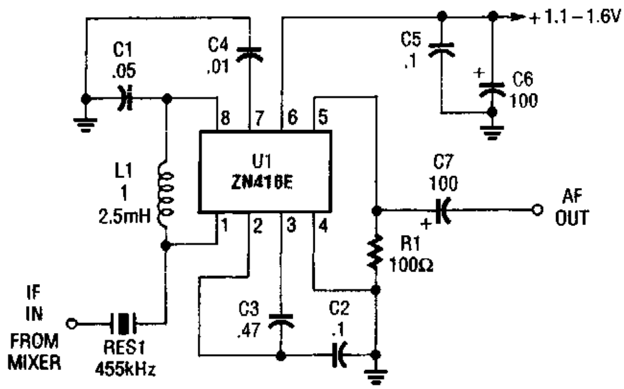
20-W 1296-MHz AMPLIFIER MODULE



73 AMATEUR RADIO *FIG. 86-15*

Using a Mitsubishi M57762 amplifier module, this amplifier delivers 20-W output on 1296 MHz. A single 12-V nominal power supply can be used.

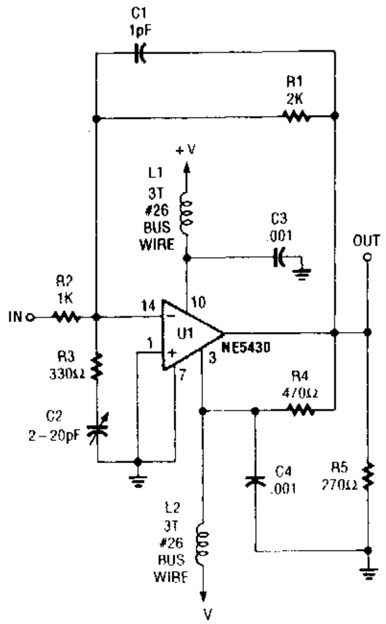
SIMPLE 455-kHz IF AMPLIFIER



POPULAR ELECTRONICS *FIG. 86-16*

The ZN416E can be configured as a simple 455-kHz IF amplifier. In this case, the circuit's center frequency and bandwidth are set by RES1 (a Murata CSB455E ceramic resonator).

UHF AMPLIFIER



NOTE
RESISTORS-¼ WATT CARBON
L1 & L2 WOUND ON FERROXCUBE VK200 09/3B
WIDEBAND THREADED CORE.

POPULAR ELECTRONICS *FIG. 86-17*

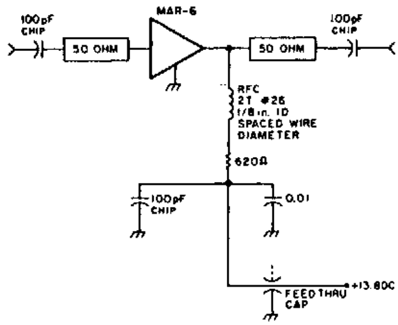
144- TO 2304-MHz UHF BROADBAND AMPLIFIER

Table 1.

Device	Max. mA	Normal Current mA.	Approx. Gain 1-GHz
MAR-1	40	20–30 mA	18 dB
MAR-2	60	30–40 mA	13 dB
MAR-3	70	30–50 mA	12 dB
MAR-4	85	50–70 mA	8 dB
MAR-6	50	15–25 mA	17 dB
MAR-7	60	25–40 mA	13 dB
MAR-8	65	30–50 mA	23 dB

Table 2.
MMIC Amplifier Performance

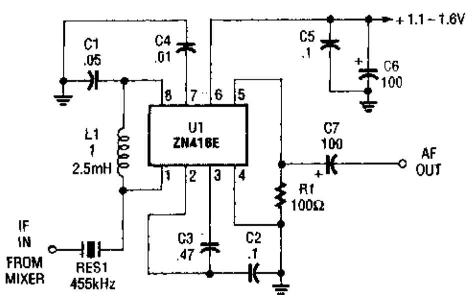
144 MHz	18.2 dB	2.7 dB N/F
220 MHz	18.3 dB	2.6 dB N/F
432 MHz	16.5 dB	2.8 dB N/F
902 MHz	15.0 dB	2.9 dB N/F
1296 MHz	13.0 dB	3.5 dB N/F
2304 MHz	8.8 dB	4.2 dB N/F



73 AMATEUR RADIO *FIG. 86-18*

Based on an MAR-6 preamp, this circuit yields low noise figures and useful gain for the 144-MHz to 2304-MHz amateur bands.

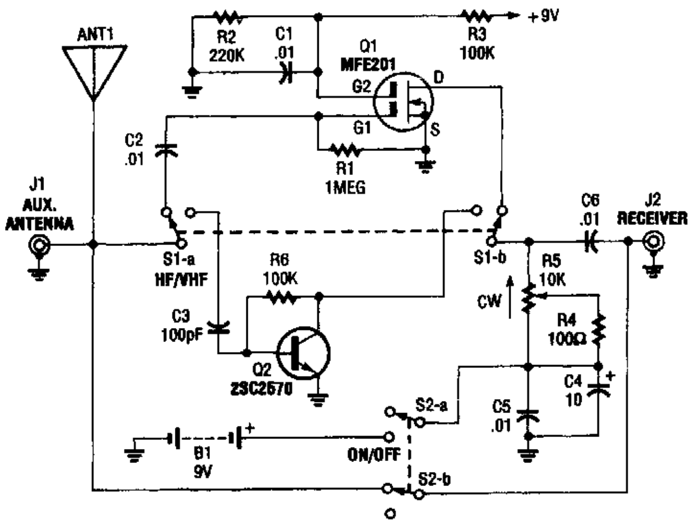
455-kHz IF AMPLIFIER



POPULAR ELECTRONICS *FIG. 86-19*

Up to 60 dB of gain at 455 kHz is available with the MC1350P. RES1 is a ceramic resonator, LC, or crystal filter. Keep the leads to pins, 1, 2, 3, and 7 short.

SWITCHABLE HF/VHF ACTIVE ANTENNA

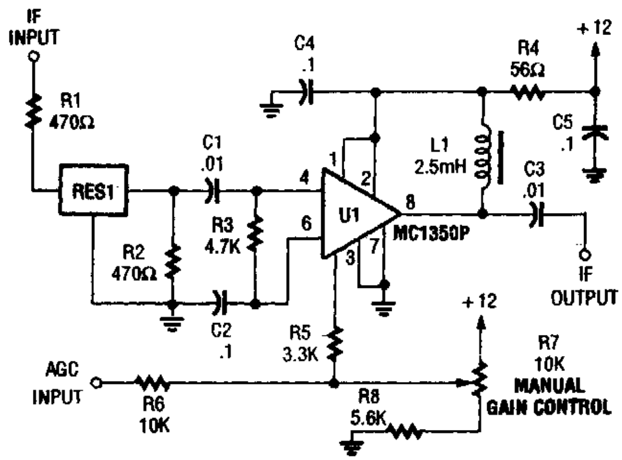


POPULAR ELECTRONICS

FIG. 86-20

The AA-7 active antenna contains only two active elements: Q1 (an MFE201 N-channel dual-gate FET) and Q2 (a 2SC2570 npn VHF silicon transistor), which provide the basis of two independent, switchable RF preamplifiers.

455-kHz IF AMP FOR 1.5-V OPERATION

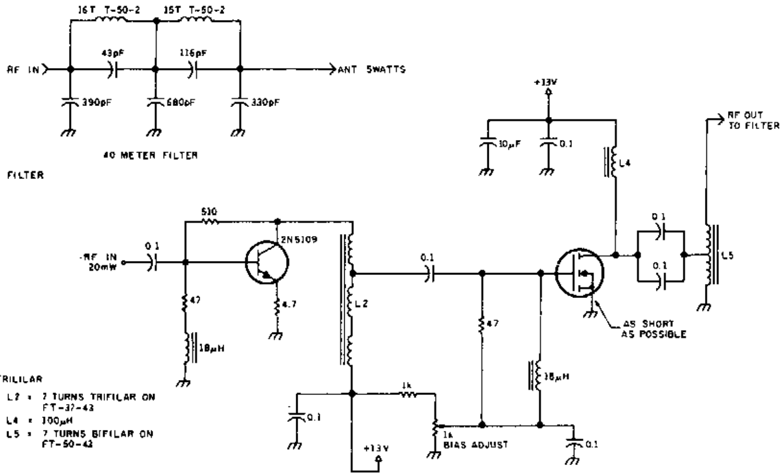


POPULAR ELECTRONICS

FIG. 86-21

The ZN416E can be configured as a simple 455-kHz IF amplifier. In this case, the circuit's center and bandwidth are set by RES1 (a Murata CSB455E ceramic resonator).

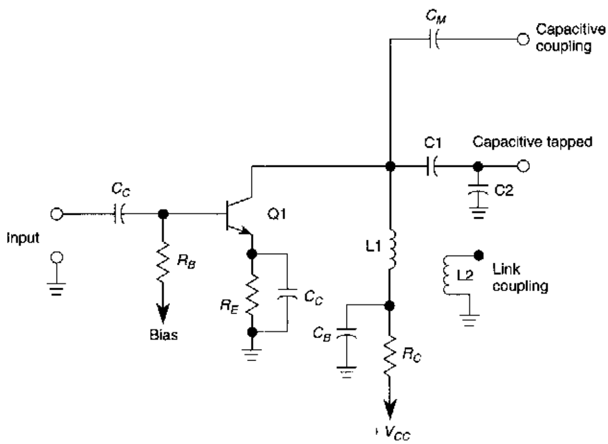
5-W 7-MHz RF POWER AMPLIFIER



73 AMATEUR RADIO TODAY ***FIG. 86-22***

The circuit shown will produce up to 5-W RF output in the 40-m (7 MHz) amateur band. The coils shown are wound on toroidal cores (Armdon Associates Inc.). The part numbers are given in the schematic. The circuit requires about 20-mW drive and a 13-V supply.

LC TUNED AMPLIFIERS



$$f_O = \frac{1}{2\pi\sqrt{LC\text{ effective}}}$$

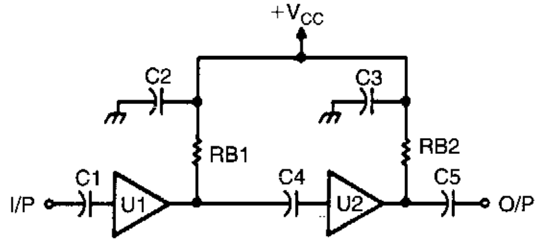
WILLIAM SHEETS ***FIG. 86-23***

This basic tuned LC amplifier can be used with three output coupling methods. They are capacitive coupling output, capacitive tapped output, or link-coupled output.

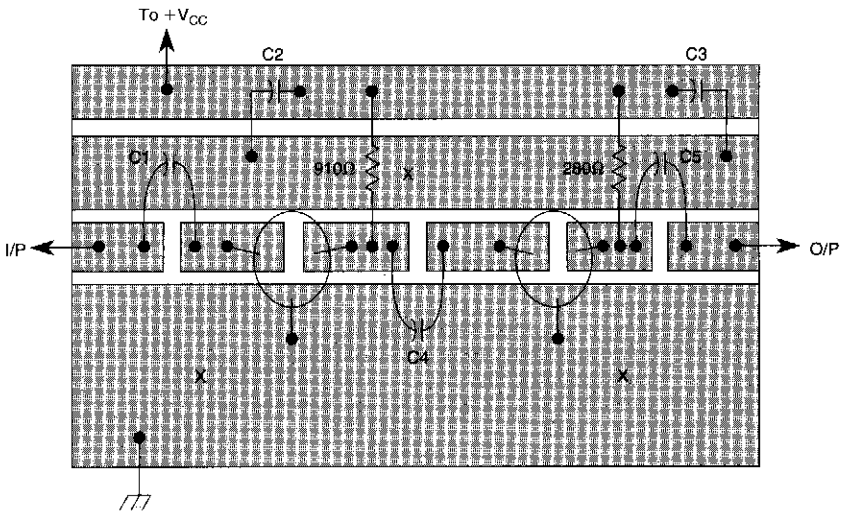
WIDEBAND PREAMP

Table 1.

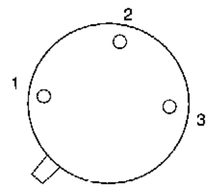
	Vcc	Vd	Id	Rb
MWA110	5 Vdc	2.9 Vdc	10 mA	210Ω
	6			310Ω
	12			910Ω
MWA120	5	5.0	25	1Ω
	6			40Ω
	12			280Ω
MWA130	5	3.2	25	85Ω
	6			120Ω
	12			360Ω



V_{CC} = 12 Vdc; C1 to C5 = 0.1μF; RB1 = 910Ω; RB2 = 280Ω; U1 = MWA110; U2 = MWA120



Pin out



1-Input

2-Output

3-Ground (case)

PC board layout (shading represents copper) and parts layout. "X" is the feedthrough wire to the gound plane. All capacitors are 0.1 μF. Keep all leads short.

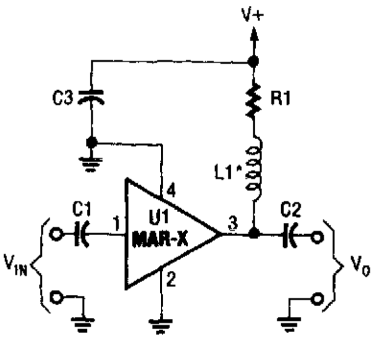
73 AMATEUR RADIO **_FIG. 86-24_**

Motorola MWA 110, 120, or 130 are wideband amplifier ICs. This wideband preamp circuit can be used in many applications. Keep the leads short when constructing the circuitry.

RF PREAMPLIFIERS

TABLE 1—MAR-X CAPABILITIES

DEVICE	MAX. FREQ. (MHz)	GAIN (100/50/1000 MHz)	N.F.	COLOR
MAR-1	1,000	18.5/17.5/15.5	5	Brown
MAR-2	2,000	13/12.8/12.5	6.5	Red
MAR-3	2,000	13/12.8/12.5	6	Orange
MAR-4	1,000	8.2/8.2/8	7	Yellow
MAR-6	2,000	20/19/16	2.8	White
MAR-7	2,000	13.5/13.1/12.5	5	Violet
MAR-8	1,000	33/28/23	3.5	Blue

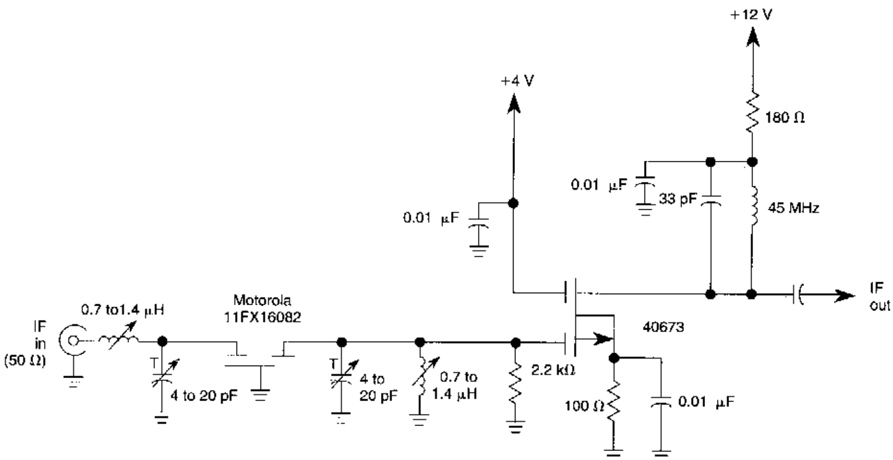


In this basic MAR-x-based circuit, both the input and output are comprised of a single dc-blocking capacitor (C1 and C2 for the input and output, respectively). The dc power-supply network (comprised of L1 and R1) is attached to the MAR-x via the RF-output terminal (lead 3).

POPULAR ELECTRONICS

FIG. 86-25

45-MHz IF AMPLIFIER WITH CRYSTAL FILTER



WILLIAM SHEETS

FIG. 86-26

A 40673 dual-gate MOSFET is matched to a crystal filter at 45 MHz. The filter impedance is around 2kΩ. The + 4-V source can be made variable for gain control (about +4 to −4V.)

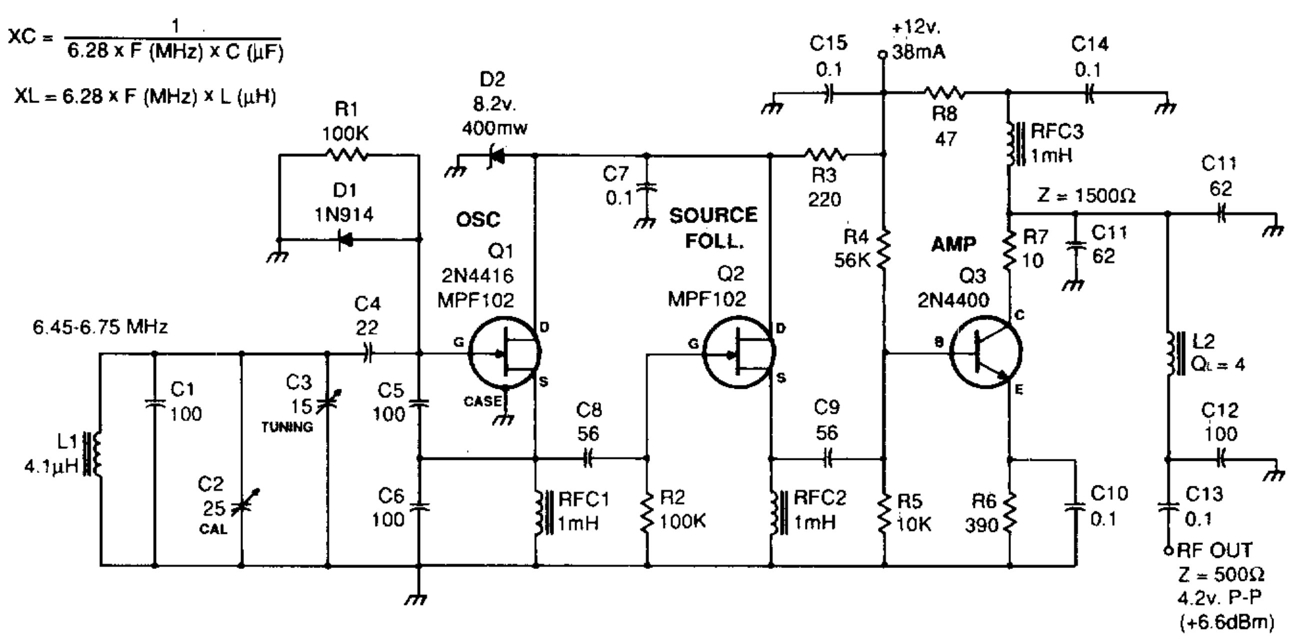
87

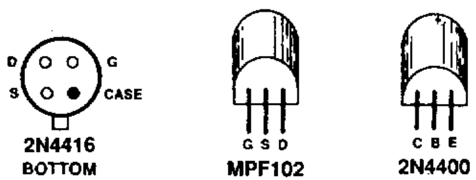
RF Oscillator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

6.5-MHz VFO
RF Signal Generator
NE602 RF Oscillator Circuits
A Shortwave Pulsed-Marker Oscillator
Ham Band VFO

6.5-MHz VFO



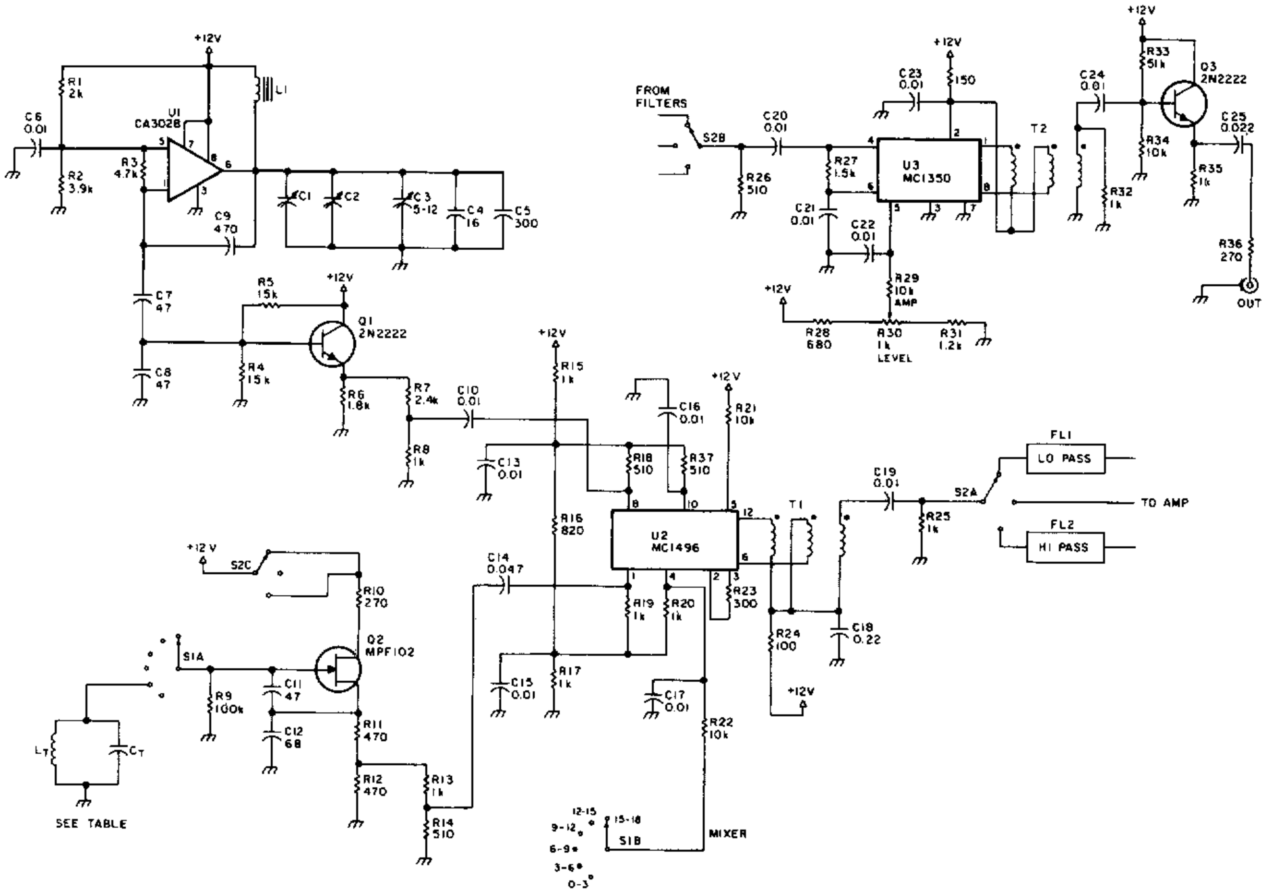


Schematic diagram of the VFO. Fixed-value capacitors are disc ceramic. C1, C4, C5, C6, and C8 are NP0 ceramic or polystyrene. C2 is a 25 pF ceramic trimmer and C3 is a 15 pF miniature air variable. Resistors are ¼ watt carbon film or composition. The RF chokes are miniature Mouser Electronics No. 43LR103 units. For L1 use 32 turns of No. 28 enamel wire on an Amidon Assoc. T50-6 (yellow) toroid. L2 has 25 turns of No. 28 enamel wire on an Amidon FT-37-61 ferrite toroid.

QST **FIG. 87-1**

Fixed-value capacitors are disc ceramics. C1, C4, C5, C6, and C8 are NPO ceramic or polystyrene. C2 is a 25-pF ceramic trimmer and C3 is a 15-pF miniature air variable capacitor. The resistors are ¼-W carbon film or composition. The RF chokes are miniature Mouser Electronics No. 43LR103 units. For L1, use 32 turns of #28 enamel wire on an Amidon Assoc. T50-6 (yellow) toroid. L2 has 25 turns of #28 enamel wire on an Amidon Ft-37-61 ferrite toroid.

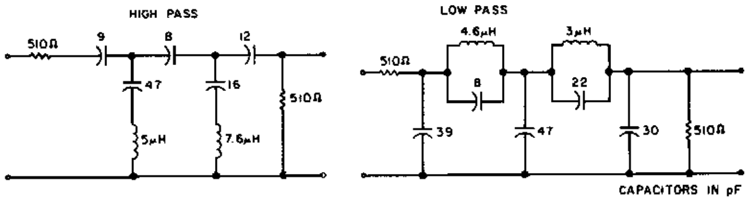
RF SIGNAL GENERATOR



73 AMATEUR RADIO TODAY

FIG. 87-2A

This circuit uses a VFO operating from 15 to 18 MHz (U1), which feeds a balanced mixer (U2). A fixed oscillator signal is mixed with this signal to generate an output from 0.4 to 33 MHz. FL1 and FL2 are low- and high-pass filters that are used to eliminate undesired mixer products. Amplifier U3/Q3 supplies up to 200 mV rms to the output jack.

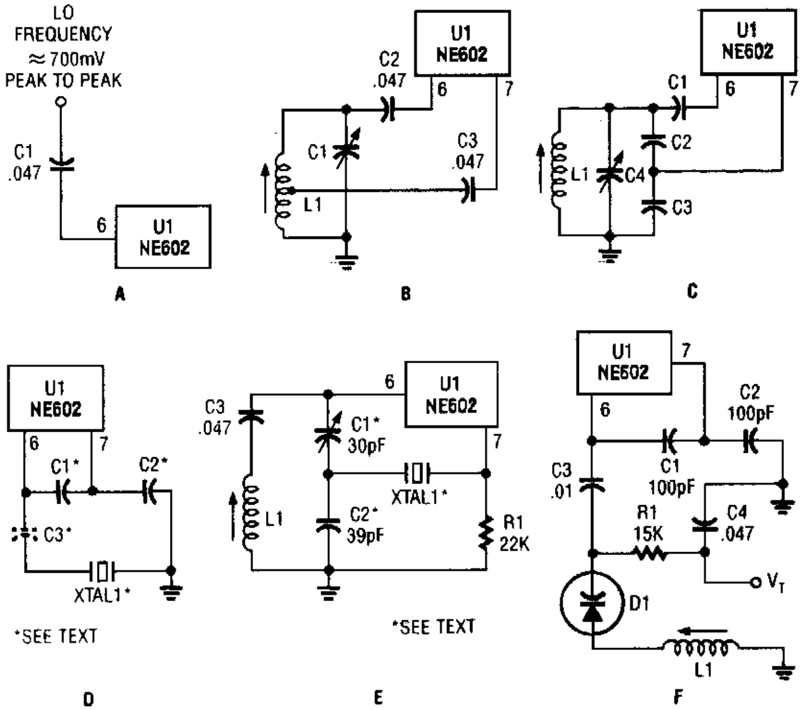


FILTER SCHEMATIC

73 AMATEUR RADIO TODAY

FIG. 87-2B

NE602 RF OSCILLATOR CIRCUITS

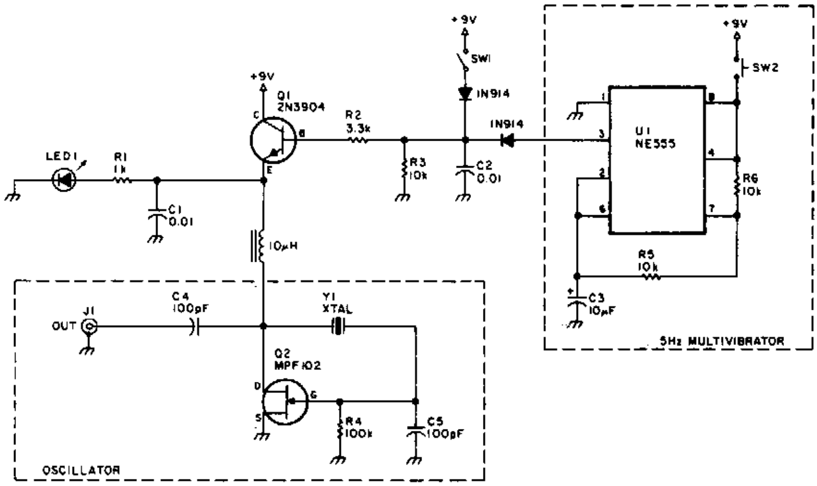


POPULAR ELECTRONICS

FIG. 87-3

Just about any standard oscillator (such as a Colpitts or Hartley configuration) can be used to generate the LO (local oscillator) frequency needed by the NE602.

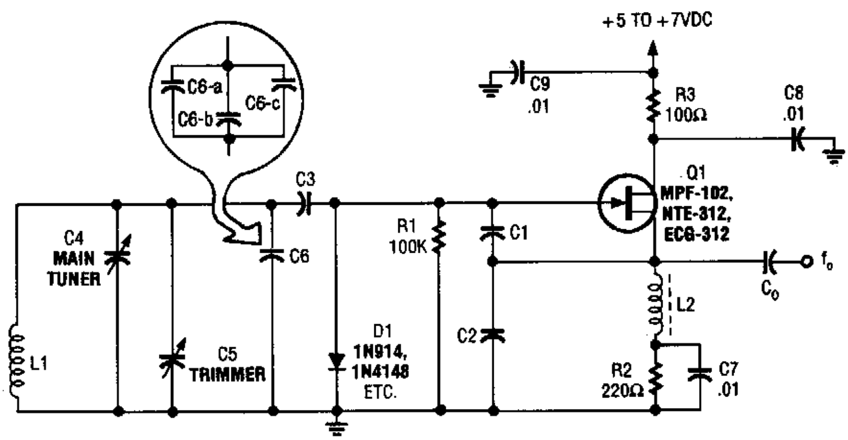
A SHORTWAVE PULSED-MARKER OSCILLATOR



73 AMATEUR RADIO TODAY ***FIG. 87-4***

A useful marker oscillator can be made using an NE555 to pulse the oscillator at an audio rate. This makes it easy to find the signal in the presence of interference. The crystal can be any suitable frequency from 1 to 30 MHz.

HAM BAND VFO



POPULAR ELECTRONICS ***FIG. 87-5***

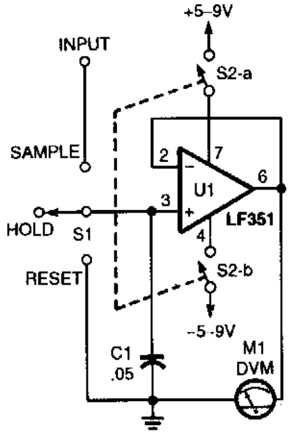
This basic VFO for the 3- to 6-MHz range is commonly used in amateur applications, using a Colpitts circuit. For 5 to 5.5 MHz, $C_1 = C_2 = 70$ pF and for 3.5 to 4.0 MHz, use 1000 pF. C3 is typically 10 to 220 pF, depending on the frequency. C4, C5, and C6, together with C3, determine the frequency along with L1. C6 can be made up of several smaller values, paralleled to get the exact required value.

88

Sample-and-Hold Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SAMPLE-AND-HOLD CIRCUIT I

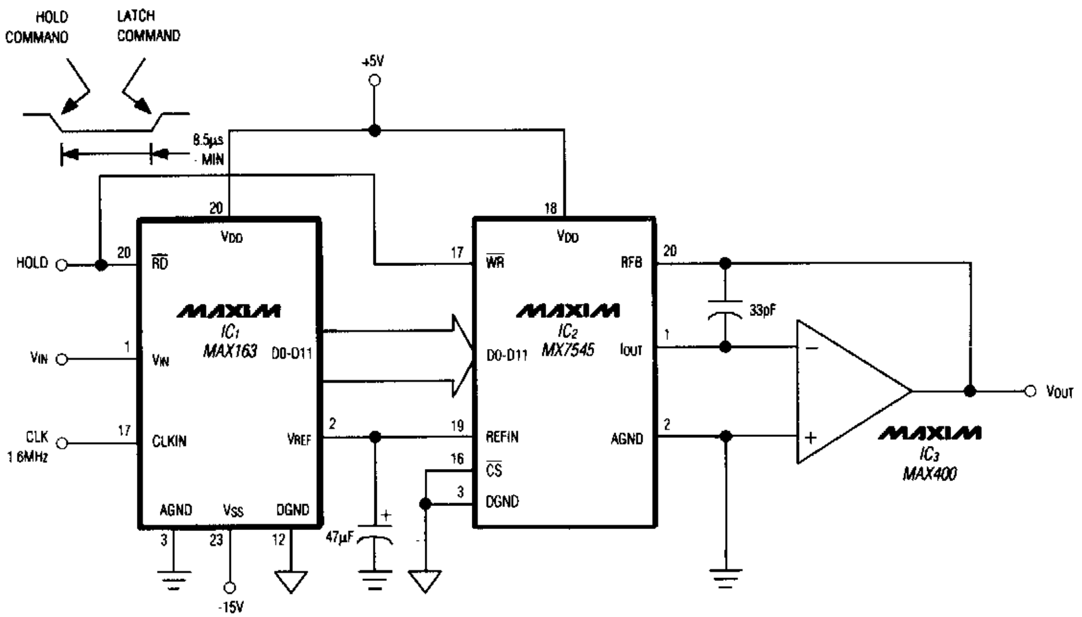


This circuit demonstrates the principle of the sample-and-hold circuit. S1 can be replaced by electronic switches (FET, etc.) in an actual application.

POPULAR ELECTRONICS

FIG. 88-1

SAMPLE-AND-HOLD CIRCUIT II



MAXIM ENGINEERING JOURNAL

FIG. 88-2

Driving a D/A converter with an A/D converter provides an overall analog-hold function, which though limited in output resolution, offers zero voltage droop and infinite hold time. The A/D converter shown (IC1) includes a 12-bit compatible track/hold at its input. The track/hold specifies a 6-MHz full-power bandwidth, a 30-ns aperture delay, and a 50-ps aperture jitter. The direct connections shown allow the D/A converter to reconstruct signal levels within the input range of 0 to 5 V.

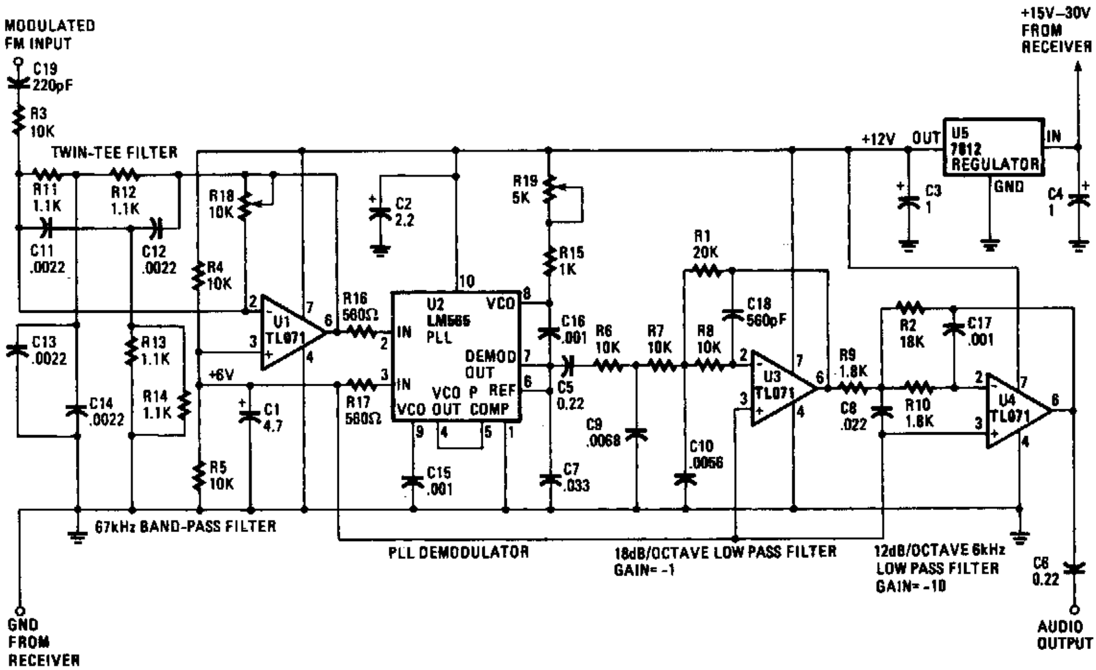
89

SCA Circuit

The source of the following circuit is contained in the Sources section, which begins on page 675. The figure number in the box of the circuit correlates to the entry in the Sources section.

Subcarrier Adapter for FM Tuners

SUBCARRIER ADAPTER FOR FM TUNERS



1990 PE HOBBYIST HANDBOOK ***FIG. 89-1***

Op amp U1 and its associated components comprise the 67-kHz bandpass filter. A twin-T network, comprised of four 1100-Ω resistors and four 0.0022-μF capacitors, is connected in the feedback network of the op amp. That gives some gain at 67 kHz and heavy attenuation for frequencies above and below that frequency.

An additional passive filter at the input to the twin-T network (containing a 220-pF capacitor and a 10,000-Ω resistor) provides some additional roll-off for frequencies below 67 kHz.

In practice, the bandpass-filter action covers a frequency range of about 10 kHz above and below the 67-kHz center frequency. Resistor R18 sets the gain of the bandpass-filter stage.

Integrated-circuit U2 is a National LM565 phase-locked loop that modulates the 67-kHz frequency-modulated (FM) signal from U1. The LM565 PLL consists of a voltage-controlled oscillator (VCO) set to 67 kHz, and a comparator that compares the incoming frequency-modulated 67-kHz signal at pin 2 with the VCO signal that is fed into pin 5.

The output of the comparator represents the phase difference between the incoming signal and the VCO signal. Therefore, the output is the audio modulated by the subcarrier. A treble deemphasis of 150 μs is provided by a 0.033-μF capacitor (at pin 7).

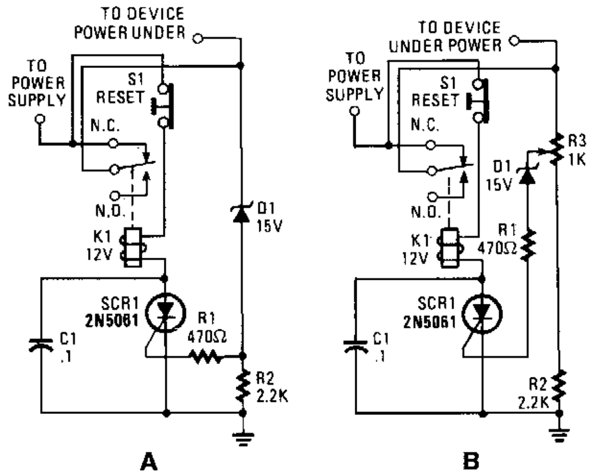
The free-running VCO frequency is determined by the 0.001-μF capacitor at pin 9 and by the resistance between the positive rail and pin 8 (100 Ω in series with R19). Variable-resistor R19 adjusts the oscillator frequency (also known as the *center frequency*) so that the incoming signal is within the lock range of the PLL.

90

Shutdown Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

RESETTABLE SHUTDOWN CIRCUITS



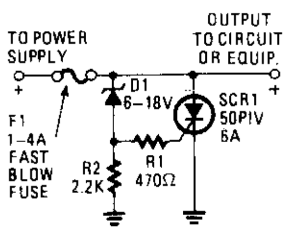
POPULAR ELECTRONICS **FIG. 90-1**

If your circuits experience frequency overvoltage conditions, continually replacing blown fuses can get pretty expensive. However, this shutdown circuit overcomes that deficiency by replacing the fuse with a relay and a low-current SCR.

When the input voltage rises above the threshold set by the Zener diode (D1), a current of sufficient magnitude is applied to the gate of SCR1, which turns it on. That draws current through the relay coil and energizes it, which swings its commutator to its normally open contact, and disrupts power to the circuit under power. Switch S1, a normally closed pushbutton switch, is used to reset the circuit; it does so by interrupting power to the relay. When S1 is pressed, the relay's wiper arm returns to the normally closed position, restoring power to the connected circuit.

If you deal with a number of circuits that have different burn-out levels, try the circuit in B. That circuit variation, a variable trip-point shutdown circuit, allows you to adjust the shutdown threshold to whatever level you desire. The circuit adjustment allows for the 30% variance in the trip point. The zener diode should be selected to have a voltage rating that is slightly lower than the minimum desired threshold voltage.

SHUTDOWN CIRCUIT



Many modern devices have shutdown circuits that are designed to remove power from the device under power when the voltage rises above a predetermined threshold. This one blows a fuse to protect the device under power.

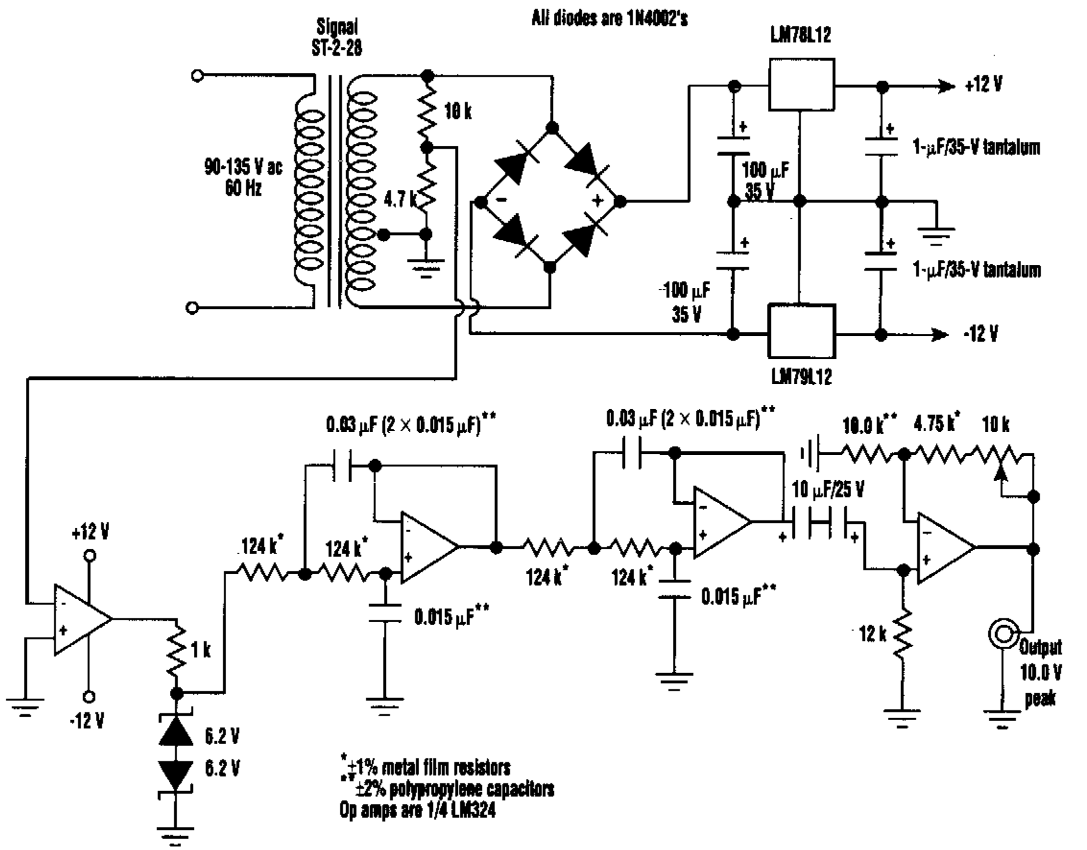
POPULAR ELECTRONICS **FIG. 90-2**

91

Sine-Wave Oscillator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

HIGHLY STABLE 60-Hz SINE-WAVE SOURCE



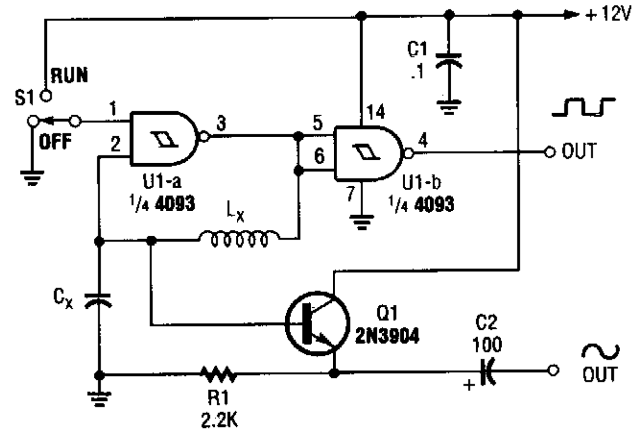
ELECTRONIC DESIGN *FIG. 91-1*

A highly-stable 60-Hz sine wave can be delivered with this circuit, which offers a different and much simpler approach to gaining a stable amplitude. Capacitor coupling the last stage removes any dc component caused by unequal zener voltages in the clipping circuit that follows the comparator.

SIMPLE SINE-WAVE OSCILLATOR

*SEE TEXT

C_x	L_x	f_{OUT}
.018	50mH	9kHz
.018	2mH	14kHz
.047	5mH	5.5kHz
1	1H	300Hz
1	10H	100Hz

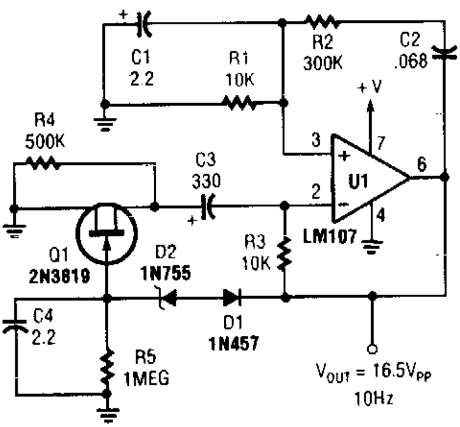


POPULAR ELECTRONICS

FIG. 91-2

Using an LC circuit, this CMOS oscillator generates sine waves.

WIEN-BRIDGE SINE-WAVE OSCILLATOR

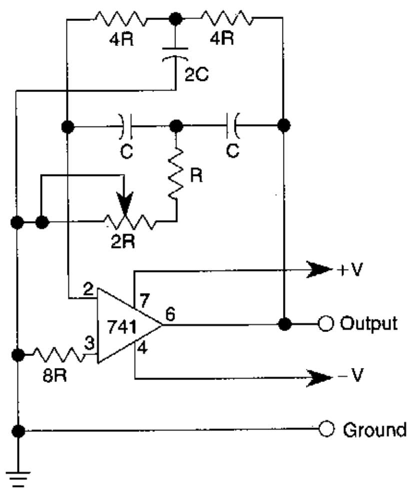


POPULAR ELECTRONICS

FIG. 91-3

This Wien-bridge sine-wave oscillator uses a 2N3819 as an amplitude stabilizer. The 2N3819 acts as a variable-resistance element in the Wien bridge.

BATTERY-POWERED SINE-WAVE GENERATOR



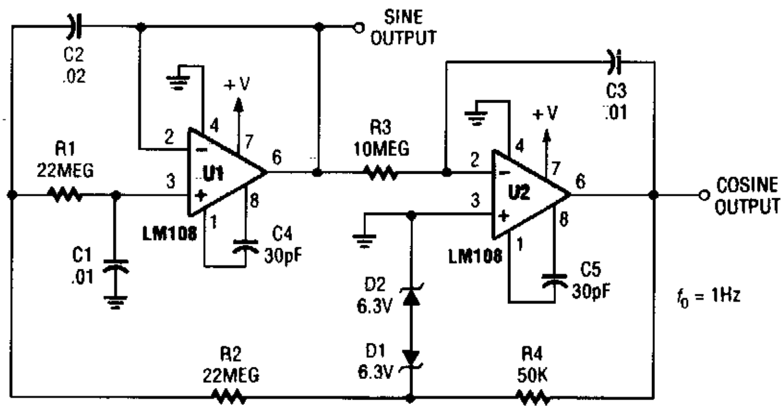
ELECTRONICS NOW

FIG. 91-4

The quality of the sine wave depends on how closely you match the components in the twin-T network in the op amp's feedback loop.

$$f = \frac{1}{2\pi RC}$$

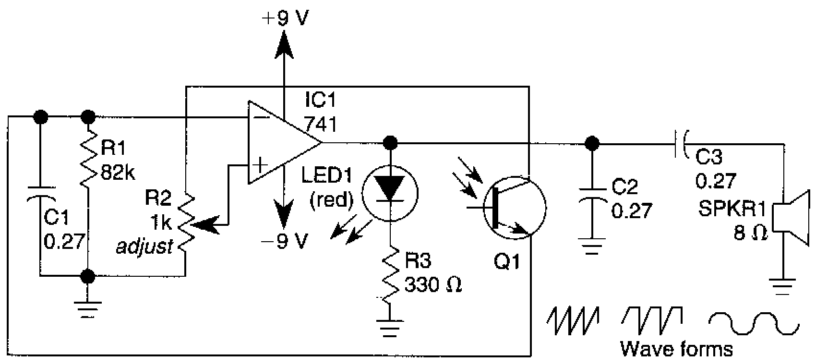
1-Hz SINE-WAVE OSCILLATOR



POPULAR ELECTRONICS *FIG. 91-5*

This circuit produces a 1-Hz sine wave using two op amps. A single-chip dual op amp could be used as well.

SIMPLE SINE-WAVE GENERATOR

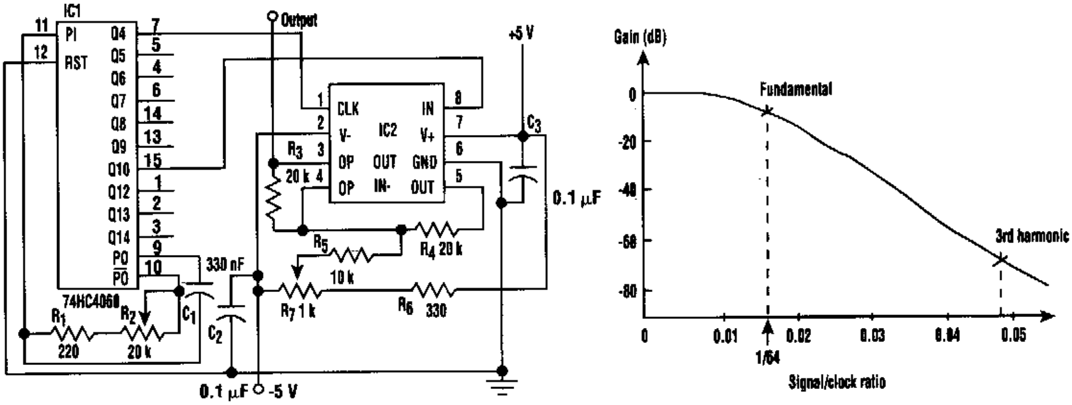


R-E EXPERIMENTERS HANDBOOK *FIG. 91-6*

A 555 timer operating in the astable mode generates the driving pulses and two 4518 dual BCD (binary coded decimal) counters provide the square waves. A TL081 op amp serves as an output buffer-amplifier, and potentiometers R1 and R2 are used in order to control the pulse's frequency and amplitude, respectively.

The output-frequency range can be varied by changing C_X. For example, a value of 0.1 µF gives a range from about 0.1 to 30 Hz, and a value of 470 pF gives a range from about 10 Hz to 1.5 kHz. The maximum output frequency is 30 kHz.

SINE-WAVE GENERATOR

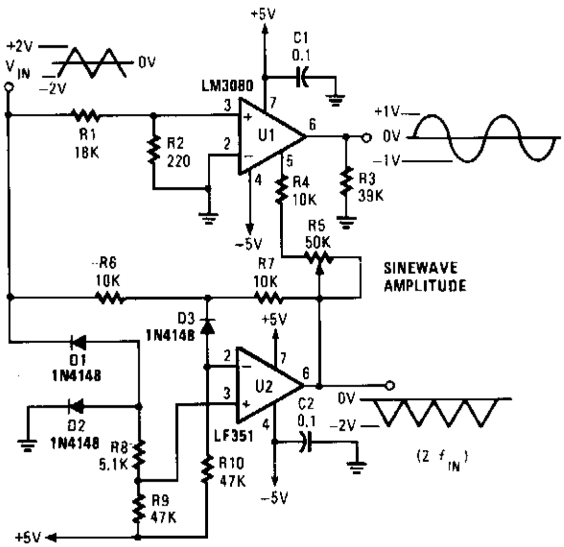


ELECTRONIC DESIGN

FIG. 91-7

In this circuit, a square wave is filtered by a high-order low-pass filter so that a –3-dB frequency will eliminate most harmonics of the waveform. As a result, the filter outputs a fundamental sine wave. This method is applied to generate a sine wave by using a switched-capacitor filter (MAX292) (see the figure). This circuit offers wide frequency range (0.1 Hz to 25 kHz), low distortion, and constant output amplitude throughout the whole frequency range.

SINE-WAVE SHAPER

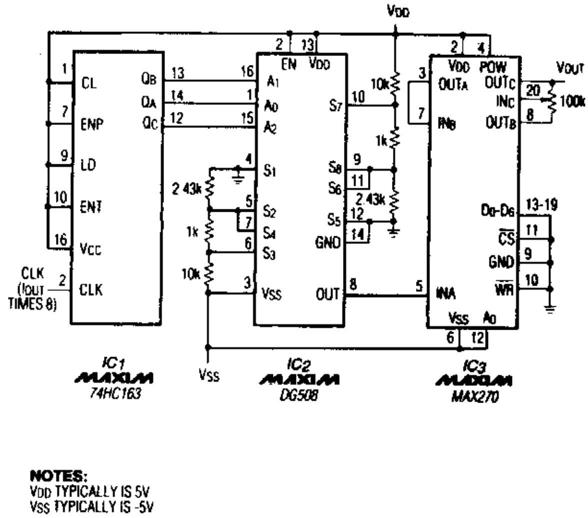


Unlike most sine-wave shapers, this circuit is temperature stable. It varies the gain of a transconductance amplifier to transform an input triangle wave into a good sine-wave approximation.

POPULAR ELECTRONICS

FIG. 91-8

PURE SINE-WAVE GENERATOR



MAXIM ENGINEERING JOURNAL ***FIG. 91-9***

A TTL counter, an 8-channel analog multiplexer, and a fourth-order low-pass filter can generate 10- to 25-kHz sine waves with a THD better than −80 dB. The circuit cascades the two second-order, continuous-time Sallen-Key filters within IC3 to implement the fourth-order low-pass filter.

To operate the circuit, choose the filter's cutoff frequency, f_C, by tying IC3's D_0 through D_6 inputs to 5 V or ground. The cutoff frequency can be at 128 possible levels between 1 and 25 kHz, depending on those seven digital input levels. Because the circuit ties D_0 through D_6 to ground, f_C equals 1 kHz. The 100-kΩ potentiometer adjusts the output level between V_{DD} − 1.5 V and V_{SS} + 1.5 V.

92

Sound- and Voice-Controlled Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

VOCAL STRIPPER

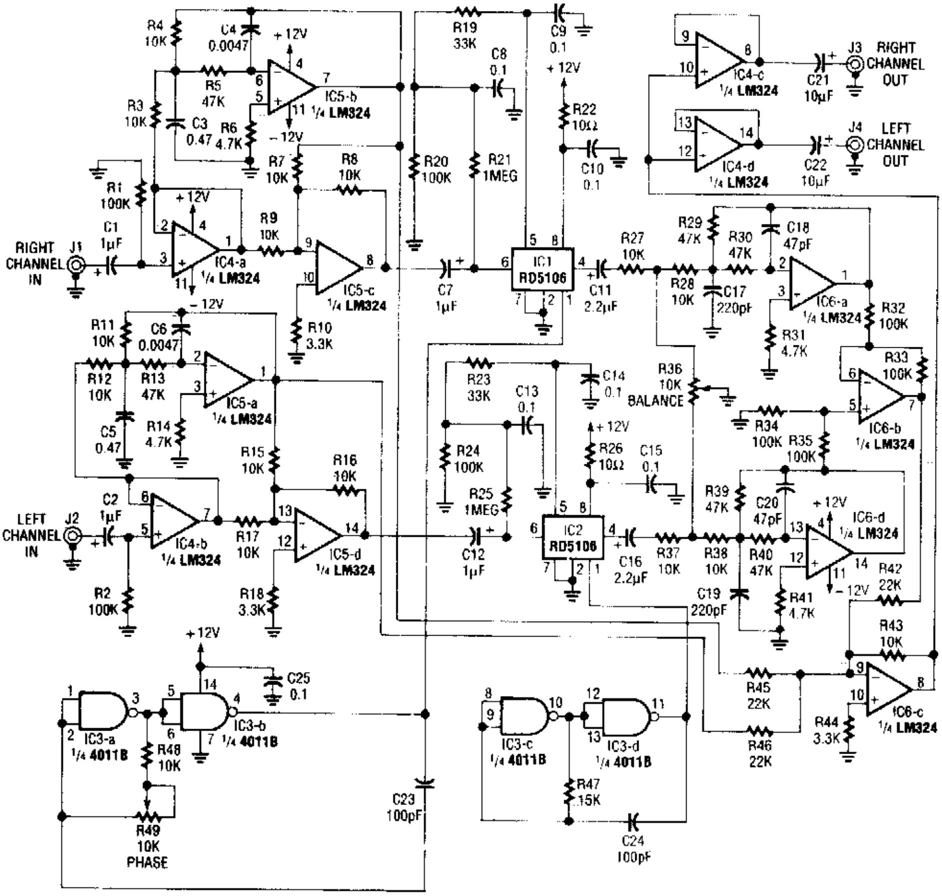


FIG. 92-1A

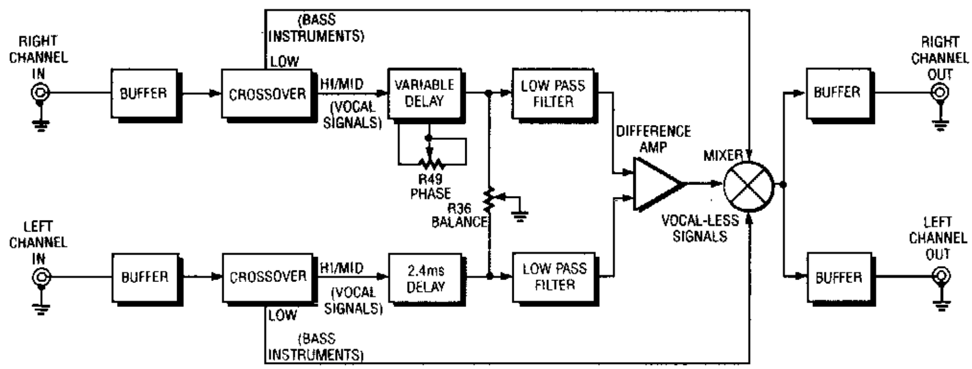


FIG. 92-1B

VOCAL STRIPPER (*Cont.*)

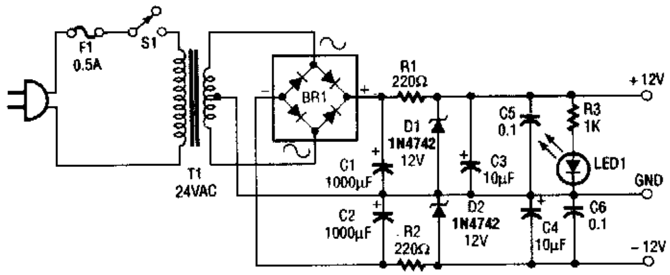
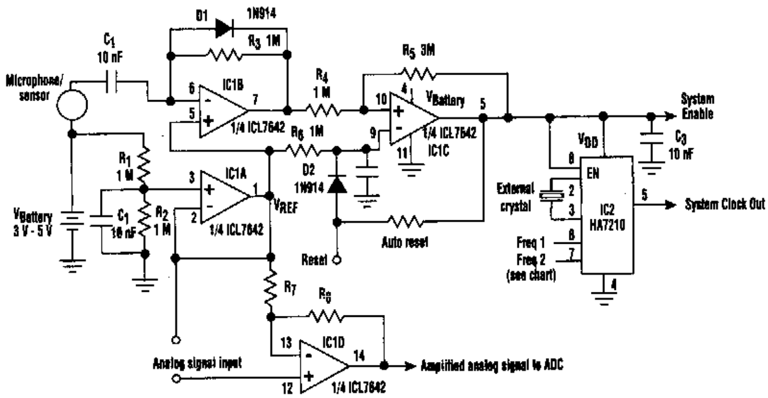


FIG. 92-1C

Right- and left-channel signals pass through 1C4-a and -b buffer amps into active crossover IC5; low frequencies are sent to the IC6-c mixer, and middle and high frequencies are sent to the analog delay lines of 1C1 and 1C2. That output passes through 1C6-a and -d to filter high-frequency sample steps. IC6-b signals are remixed with low frequencies by IC6-c and are sent to final out via IC4-c and -d buffers.

One channel (R) is a variable-delay circuit, using an analog bucket-brigade device and a variable clock frequency. This is compared in amplitude and phase to the L channel (fixed delay). The local can therefore be nulled out via R36.

SLEEP-MODE CIRCUIT



HA7210 OSCILLATOR CONTROL INPUTS

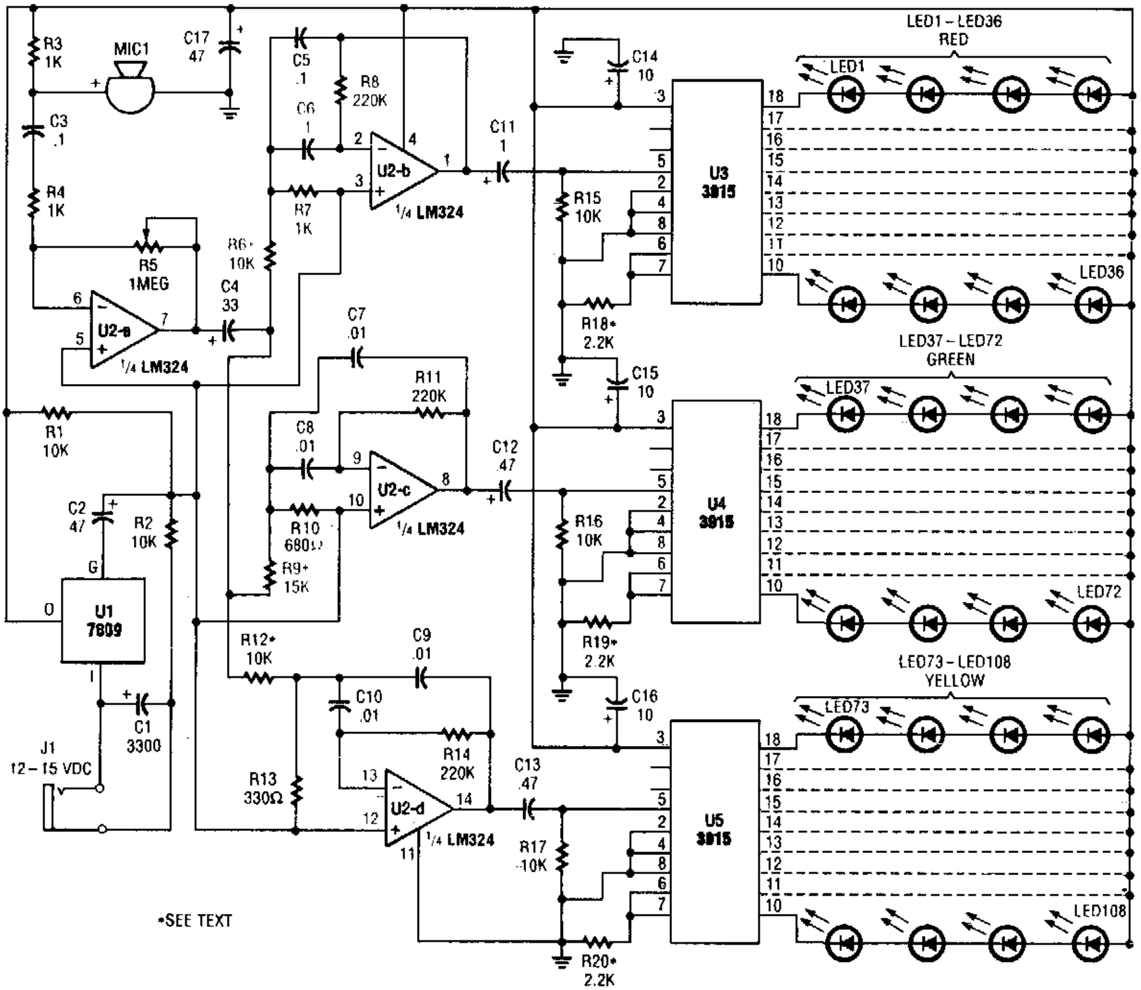
Enable	Freq 1	Freq 2	Output range
1	1	1	10 kHz-100 kHz
1	1	0	100 kHz-1 MHz
1	0	1	1 MHz-5 MHz
1	0	0	5 MHz-10 MHz +
0	X	X	High impedance

ELECTRONIC DESIGN

FIG. 92-2

The HA7210 oscillator IC combines with an ICL7642 quad CMOS op amp to produce a sleep-mode control circuit. The circuit is put into the sleep mode with a logic high applied to the Reset input or with an RC timer for automatic reset. The system is awakened by a signal from the microphone/sensor.

SONIC KALEIDOSCOPE



POPULAR ELECTRONICS ***FIG. 92-3***

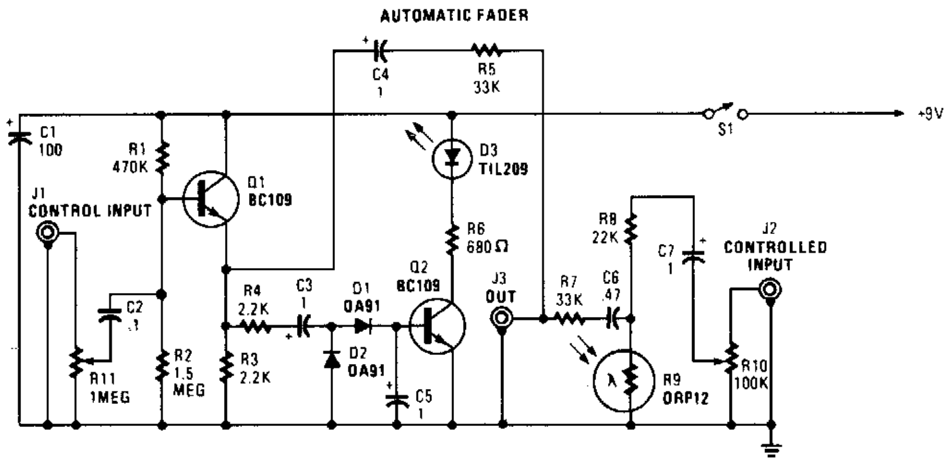
The microphone input, MIC1, is fed through C3 and R4 to inverting amplifier U2-a; the gain of U2-a is controlled by potentiometer R5. The output of U2-a is fed through C4 to the remaining op-amps (U2-b, U2-c, U2-d), which are all configured as band-pass filters. Each filter is tuned to pass a different range of frequencies by its resistor/capacitor combination. With the values shown, U2-b, U2-c, and U2-d have center frequencies of roughly 100, 1000 and 1500 Hz, respectively.

Resistors R6, R9, R12 control the bandwidth and gain of their respective filter circuits, and can range in value from 10 to 15 kΩ. The output of U2-b is capacitively coupled via C11 to the input of U3, with R15 serving as the load resistor for U2-b. That resistor also keeps U3's outputs from "floating" in the absence of a signal. Connected as shown, U3 uses its own internal voltage reference to make a full-scale display of 1.2 V.

Each of the nine outputs of U3 (output 1 is not used) sinks four, series-connected (red) LEDs. Op amps U2-c and U2-d are similarly connected to U4 and U5, respectively, driving green and yellow LED strings. Resistors R18, R19, and R20 control the brightness of their corresponding LED arrays, and they must be adjusted accordingly; different colors of LEDs usually vary in brightness. A lower value of resistance will make the LEDs glow brighter.

Power for the circuit is supplied by a 500 mA, 12–15-Vdc wall-pack transformer, via J1. The output of the transformer is filtered by C1 and is regulated by U1; regulation is necessary to keep power-line ripple from affecting the display. The supply pins of U2 through U5 are bypassed by capacitors C14 through C17 to further ensure stability. An on/off switch was deemed unnecessary because the power supply should be unplugged when the unit is not in use.

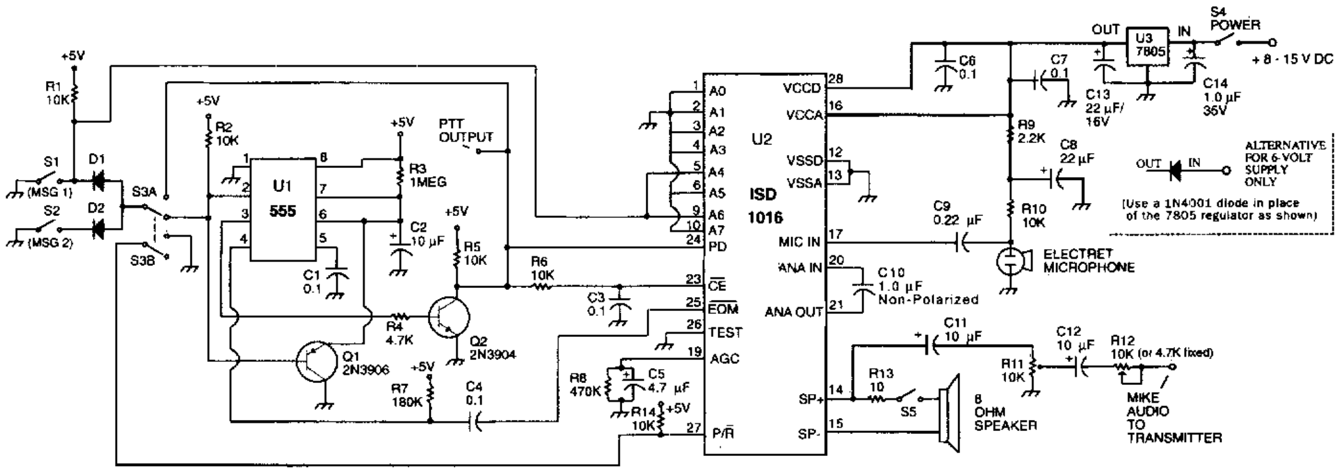
AUTOMATIC FADER



POPULAR ELECTRONICS **FIG. 92-4**

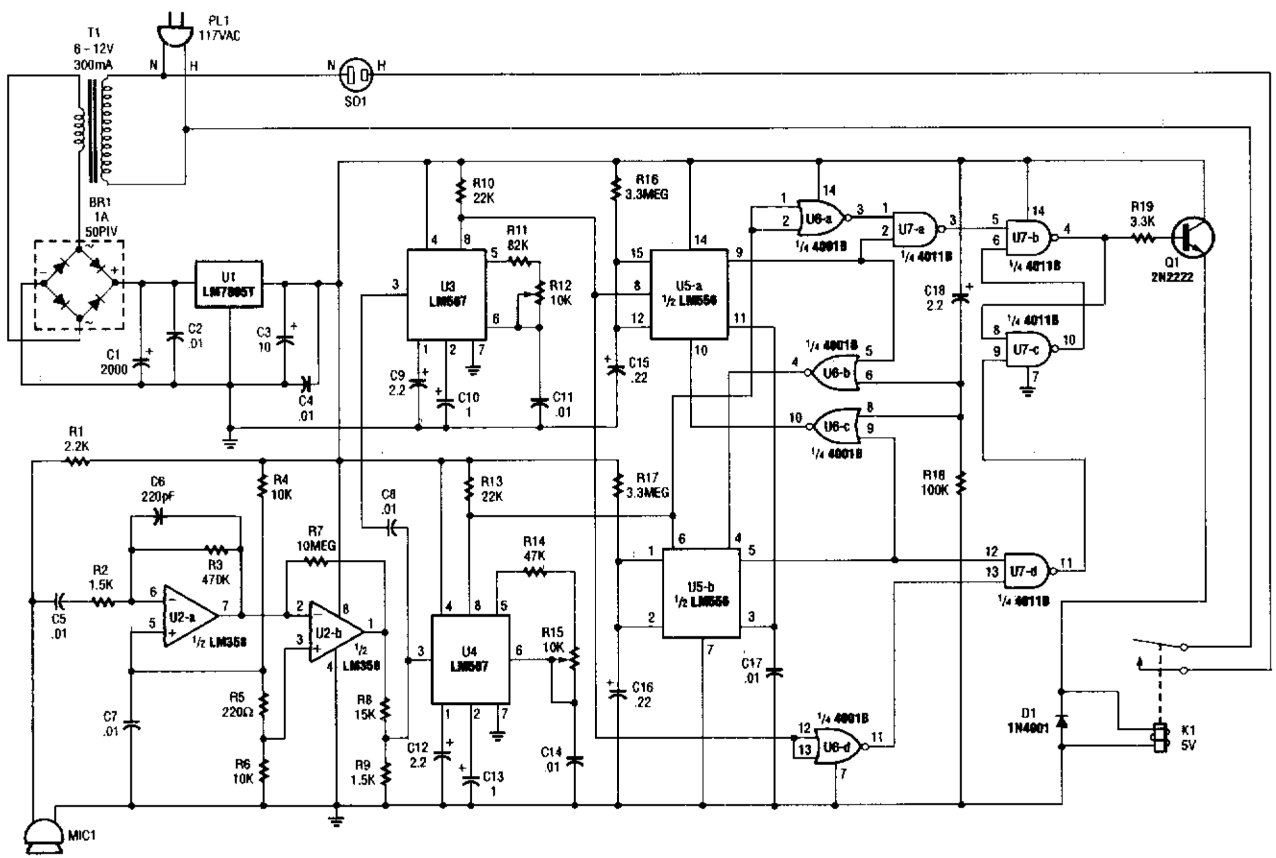
In this circuit, audio fed to the control channel is amplified and rectified by D1 and D2. This dc level activates LED D3 via Q2. The light from D3 causes R9, a light-dependent resistor to decrease resistance. As R11 (audio gain) is set higher, more audio is present at the output of Q1. Audio fed into J2 is shunted to ground via R9 and less of this audio appears at J3. Therefore, audio at J1 controls the audio level fed to J3 from J2 and produces a fade effect.

VOICE IDENTIFIER FOR HAM RADIO USE



73 AMATEUR RADIO TODAY *FIG. 92-5*

Using an ISD1016 audio record/playback chip (Information Storage Devices, Inc.), this circuit records and plays back messages on command. Although intended for use with transmitters, it can be used as an electronic notepad, etc. Consult the ISD1016 data sheet for other applications.

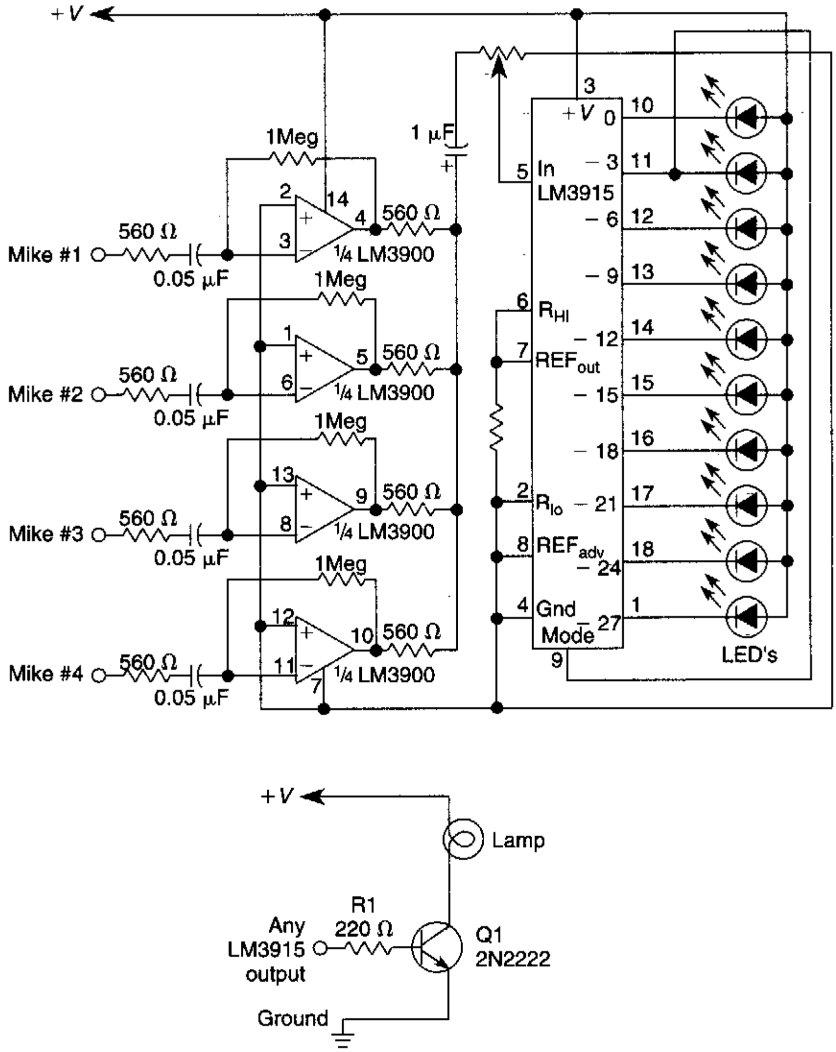


POPULAR ELECTRONICS

FIG. 92-6

At the heart of the whistle switch are a pair of tone detectors, each of which is built around an LM567 tone decoder, which are supported by a minimum of additional components. This whistle switch is designed to respond to only two or more occurrences of a specific tone, or sequence of tones, within a specified period to prevent false triggering. Depending on the relay used, various ac loads can be controlled. Microphone MIC1 picks up the sound and U2 amplifies the signal and feeds it to tone decoders U3 and U4. These devices trigger U5-a and U5-b and the logic circuits that drive relay K1.

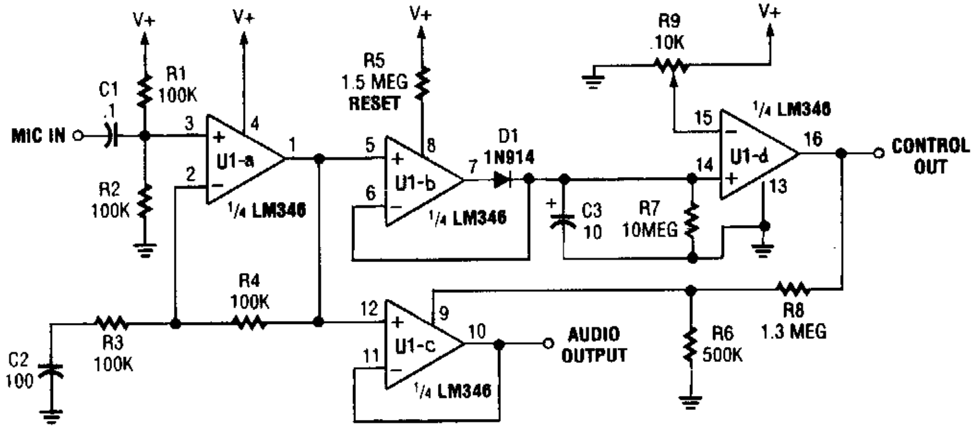
AUDIO LIGHT



RADIO ELECTRONICS *FIG. 92-7*

This circuit will produce an output when the sound exceeds a preset level. The LM3915 is a log-output bar graph driver. Use the transistor driver shown for higher current loads. To drive heavy-current loads with an LM3915 output, you must add a transistor, as shown in B.

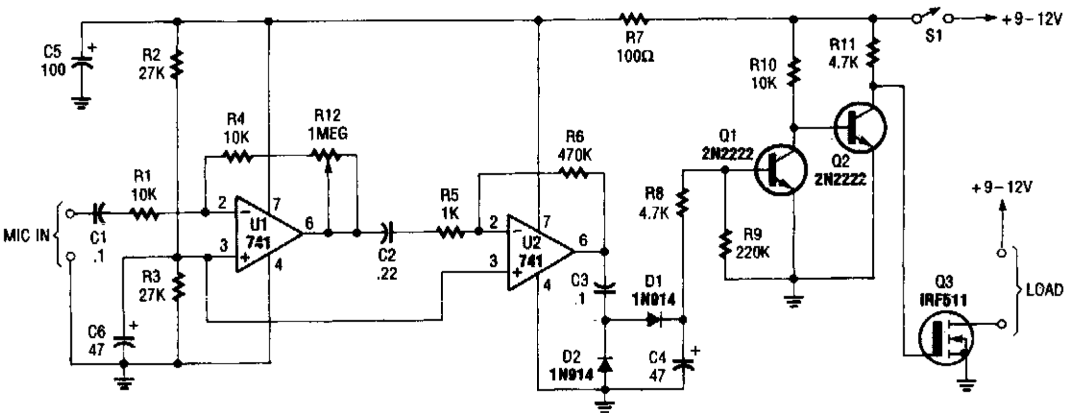
VOICE-ACTIVATED SWITCH AND AMPLIFIER



POPULAR ELECTRONICS ***FIG. 92-8***

In certain applications, such as transmitter or other communications and control applications, this circuit should be useful. Both audio output and dc control outputs are provided. R9 sets the control threshold.

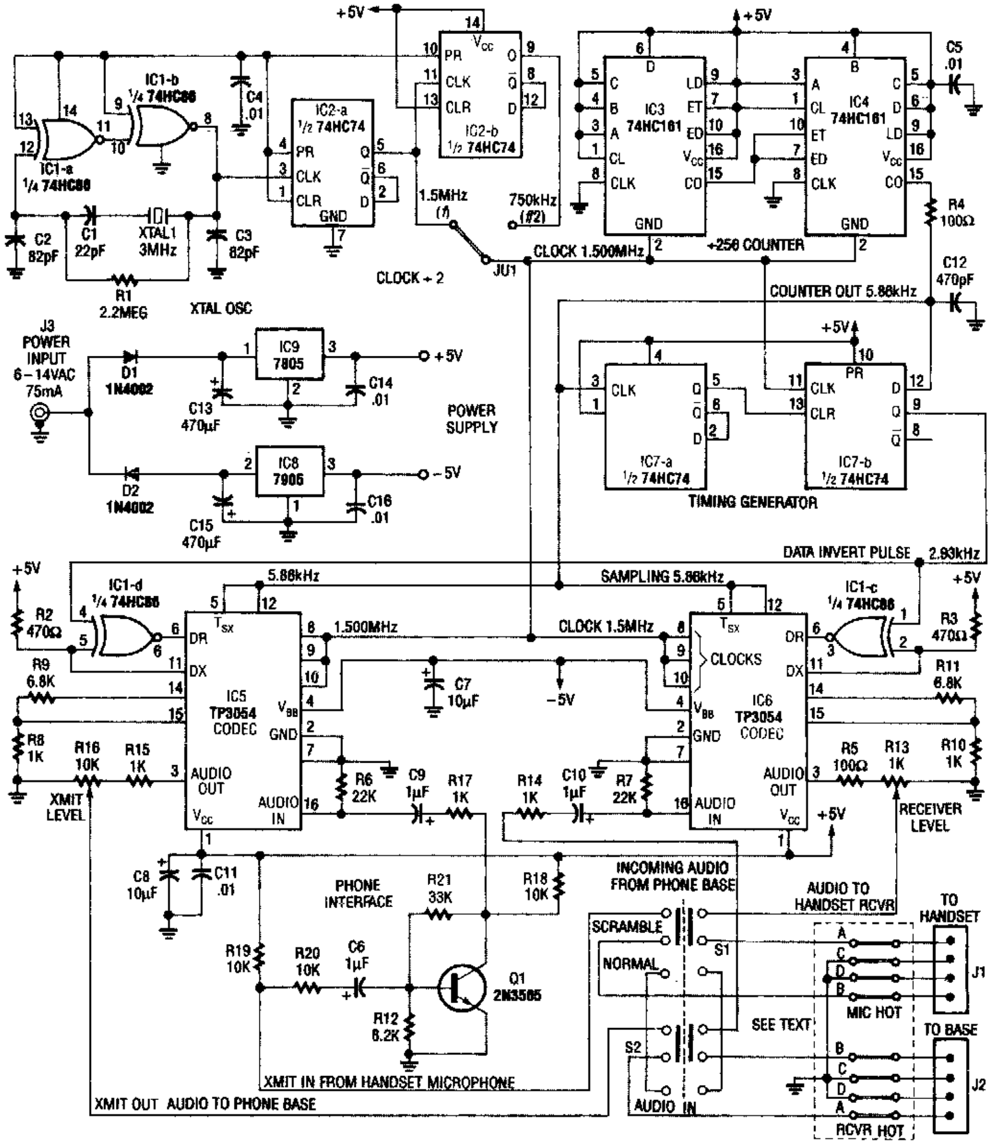
AUDIO-CONTROLLED SWITCH



POPULAR ELECTRONICS ***FIG. 92-9***

The audio-controlled switch combines a pair of 741 op amps, two 2N2222 general-purpose transistors, a hexFET, and a few support components to a circuit that can be used to turn on a tape recorder, a transmitter, or just about anything that uses sound.

SPEECH SCRAMBLER

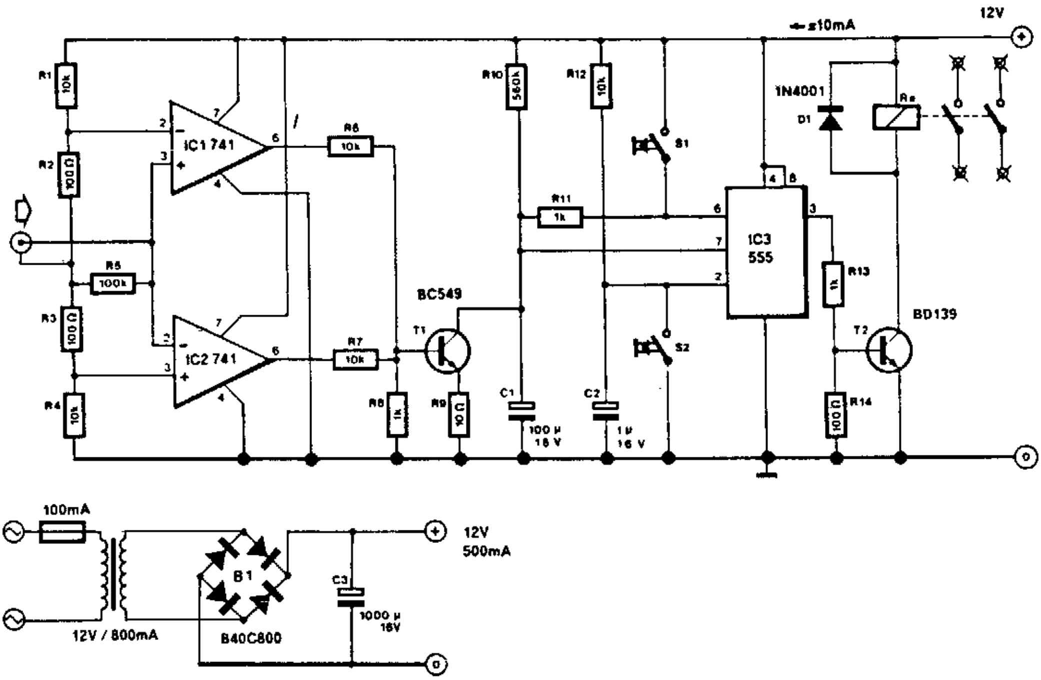


ELECTRONICS NOW *FIG. 92-10*

Using digital techniques, this circuit accomplishes the frequency-inversion algorithm via digitization of the audio, inversion of the sign of every alternate sample, and D/A conversion of the resultant data. The result is an inverted frequency spectrum. Because the circuit has two channels, this system can be used in a full duplex two-way telephone scrambler.

A complete kit of parts is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

AUDIO-CONTROLLED MAINS SWITCH



303 CIRCUITS

FIG. 92-11

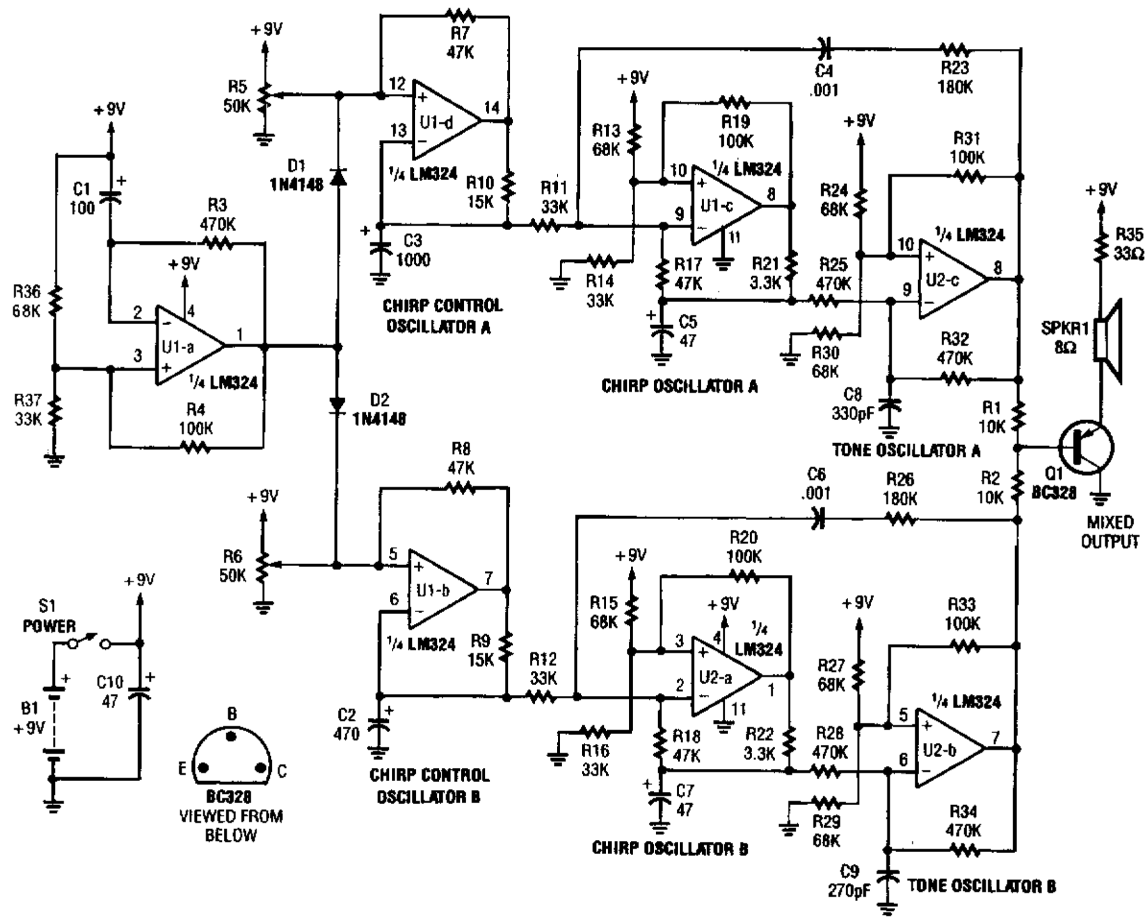
This circuit will switch off the line supply to audio or video equipment if there has been no input signal for about 2 seconds. S1 provides manual operation and S2 acts as a reset. This circuit allows for time to change a tape or compact disc. About 50 mV of audio signal is necessary.

93

Sound-Effects Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

CANARY SOUND SIMULATOR

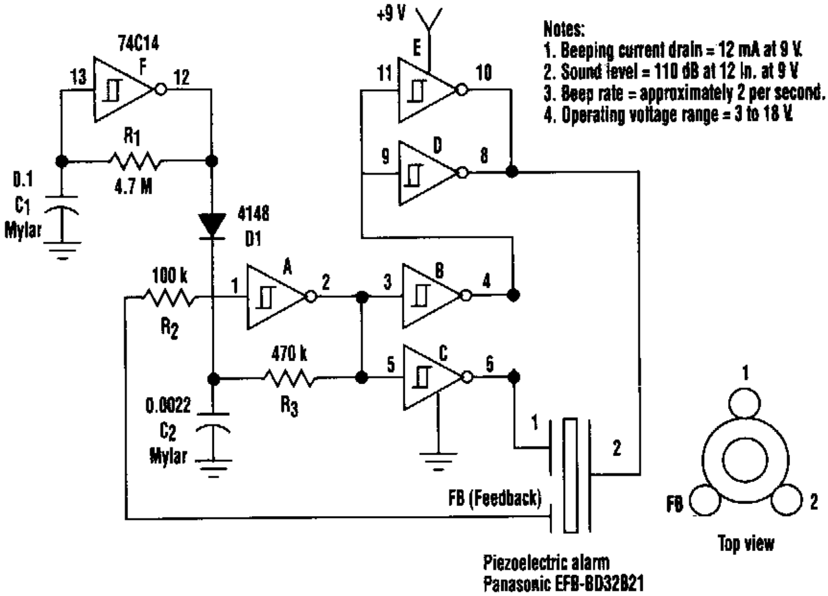


POPULAR ELECTRONICS

FIG. 93-1

This circuit generates the sound of two canaries singing in a cage. Two LM324 quad amps make up seven oscillators. One oscillator is an on/off control, the other six generate the sounds of two canaries. A 9-V supply powers the circuit.

110-dB BEEPER



ELECTRONIC DESIGN

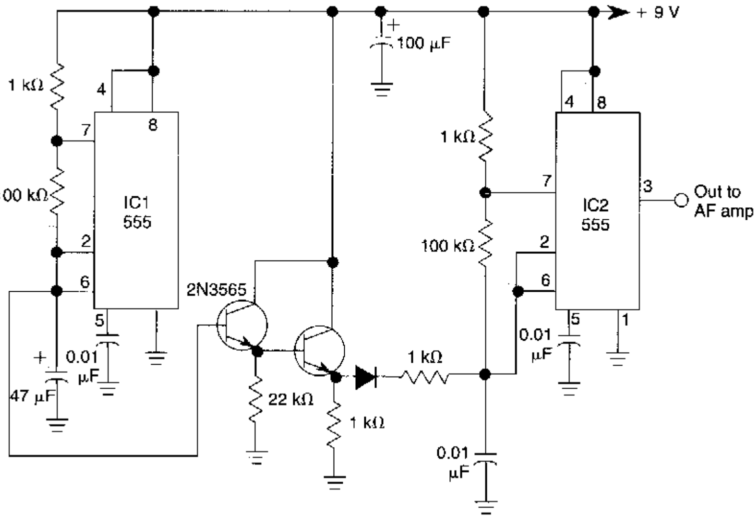
FIG. 93-2

This circuit will generate an ear-splitting 110 dB from 9 V. The setup uses a single 74C14 (CD40106B) CMOS hex inverting Schmitt-trigger IC, which must be used with a piezoelectric device with a feedback terminal. The feedback terminal is attached to a central region on the piezoelectric wafer. When the beeper is driven at resonance, the feedback signal peaks.

One inverter of the 74C14 is wired as an astable oscillator. The frequency is chosen to be 5 times lower than the 3.2 kHz resonant frequency of the piezoelectric device. Feedback from the third pin of the beeper reinforces the correct drive frequency to ensure maximum sound output.

Four other inverter sections of the IC are wired to form two separate drivers. The output of one section is cross-wired to the input of the second section. The differential drive signal that results produces about 18-V p-p when measured across the beeper. The last inverter section is wired as a second astable oscillator with a frequency of about 2 Hz. It gates the main oscillator on and off through a diode. For a continuous tone, the modulation circuit can be deleted.

SIREN ALARM

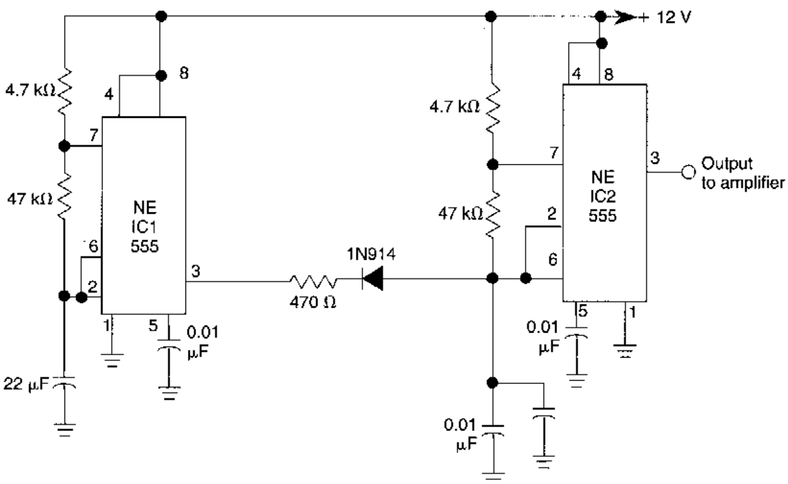


WILLIAM SHEETS

FIG. 93-3

The ramp voltage from the low frequency oscillator IC1 modulates IC2 thereby producing a rising and falling tone like the siren wail of police cars.

1000-Hz PULSED-TONE ALARM

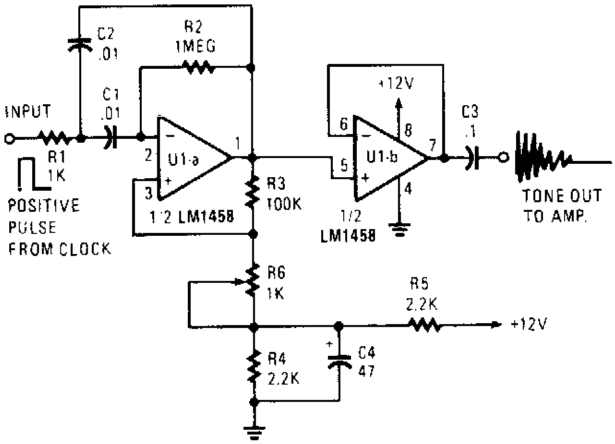


WILLIAM SHEETS

FIG. 93-4

IC1 generates a pulse that modulates the 1000-Hz tone generated by IC2. This circuit can be used to generate warning or alert signals.

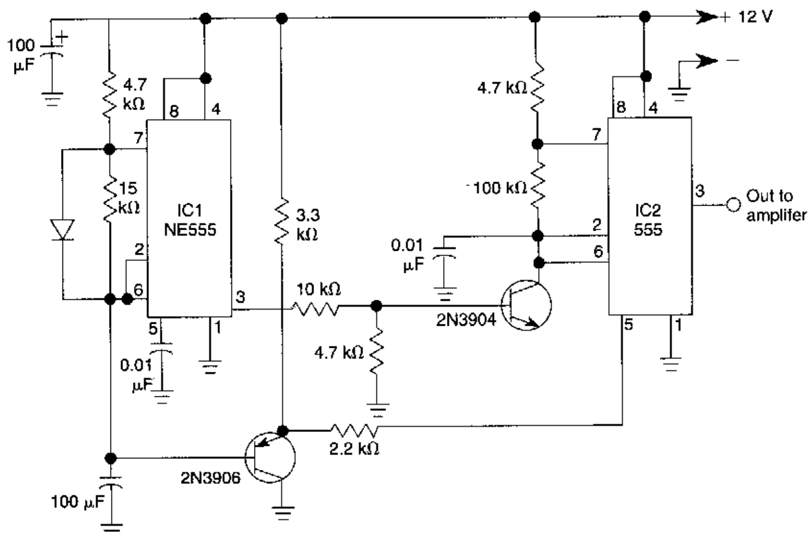
TONE CHIME



POPULAR ELECTRONICS ***FIG. 93-5***

A positive pulse input to R1 causes the active filter U1-a to "ring." If the gain is set too high (R6), the circuit will oscillate. R6 controls the positive feedback and the Q of the circuit. C1 and C2 can be changed to adjust the tone frequency.

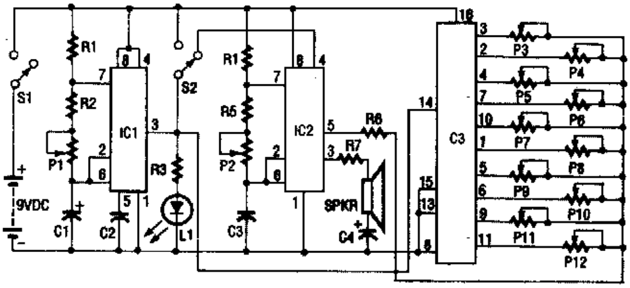
SPACESHIP ALARM



WILLIAM SHEETS ***FIG. 93-6***

By using two 555 timers this circuit produces a low frequency tone that rises to a high frequency tone in a little over 1 second. Then the sound stops for about 0.3 seconds, thereafter the cycle repeats. To produce the alarm sound of the Star Trek spaceship.

10-NOTE SOUND SYNTHESIZER



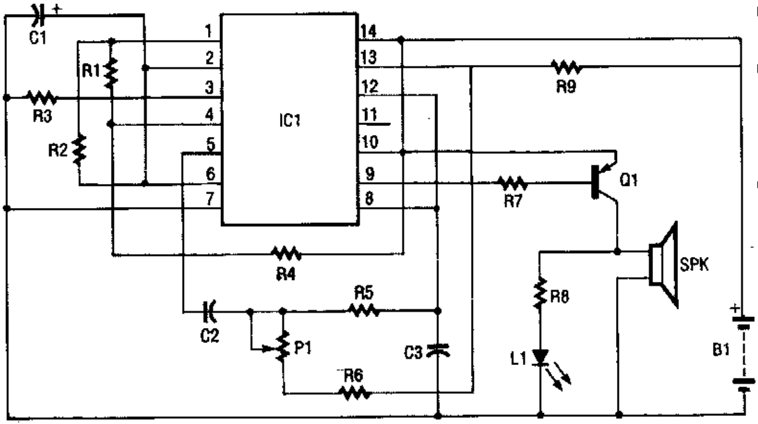
C1	22 μF Electrolytic Capacitor
C2	0.01 μF Capacitor
C3	0.1 μF Capacitor (104k)
C4	10 μF Capacitor
IC1, IC2	555 Timer IC
IC3	4017 IC
L1	Red LED
P1-P12	5k Trimmer Resistor
R1, R3,R4	1k Resistor
R2	200-ohm Resistor
R5	4.7k Resistor
R6	100-ohm Resistor
R7	15-ohm Resistor
S1, S2	SPDT Switch

1991 PE HOBBYIST HANDBOOK ***FIG. 93-7***

As shown, three ICs are used to produce the sounds. IC1 is a 555 timer that generates clock pulses. It is configured as an astable multivibrator. The frequency of the clock pulses is set by trimmer potentiometer P1. These clock pulses are coupled to the input of IC3 (a 4017 CMOS Johnson counter) on its clock input pin 14. Each clock pulse causes IC3 to shift a "high" to each of its output pins in sequence. A trimmer resistor, which can be adjusted to set a different frequency for each note, is connected to each of IC3's output pins. One side of each of the trimmers is connected to pin 5 (the control voltage pin) of IC2.

IC2, another 555 timer IC, creates the tone; the overall pitch of the tone can be varied by P2. As the output sequences from the 4017, that tone, which is changed in frequency by each output shift is applied to a small speaker from pin 3 of IC2. An LED, which flashes with each clock pulse, is connected to pin 3 of IC1. Switch S2 is used to vary the sound between "flowing" and distinct notes.

SPACE-AGE SOUND MACHINE



C1 .... 1 µF Radial Electrolytic Capacitor
C2, C3 .. 0.01 µF Disc Capacitor
IC1 .... 556 Dual Timer IC
L1 .... Green LED
P1 .... 2 Meg Trimmer Resistor
Q1 .... B714 PNP Transistor
R1 .... 82K Resistor
R2, R5 .. 33K Resistor
R3,R7 .. 4.7K Resistor
R4,R6,R9...1K Resistor
R8 .... 100 ohm Resistor
Spk ...... Small Speaker

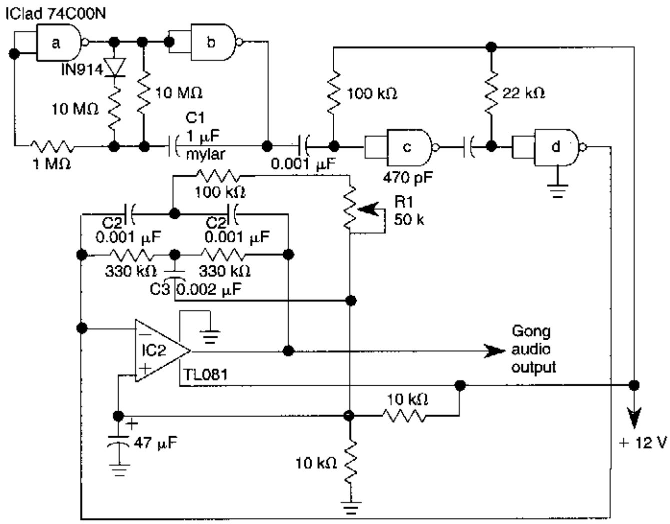
PE HOBBYIST HANDBOOK ***FIG. 93-8***

The space-age sound device uses a 556 dual-times IC to produce a phasor sound. That IC is actually two 555 timer ICs in one 14-pin package, as shown in the schematic. Each timer inside the 556 is connected in an astable multivibrator mode.

The first timer has its frequency set by R1, R2, and C1. Its output appears on pin 5 and it is coupled through C2 and R5 into the trigger input of the second timer. The second timer has an adjustable frequency that is controlled by P1, R6, and C3.

In the second timer, the first frequency mixes with the second frequency and produces the phasor-like sounds. The output of the second timer, which has the two signals mixed together, is brought from pin 9 through limiting resistor R7 to the input of Q1. The function of pnp germanium power transistor Q1 is to amplify the signal to the level that is needed to drive the speaker. The green LED, L1, converts electrons directly into visible photons (light) in time with the pulses from the speaker. The purpose of resistor R8 is to limit the current through the LED to a safe level.

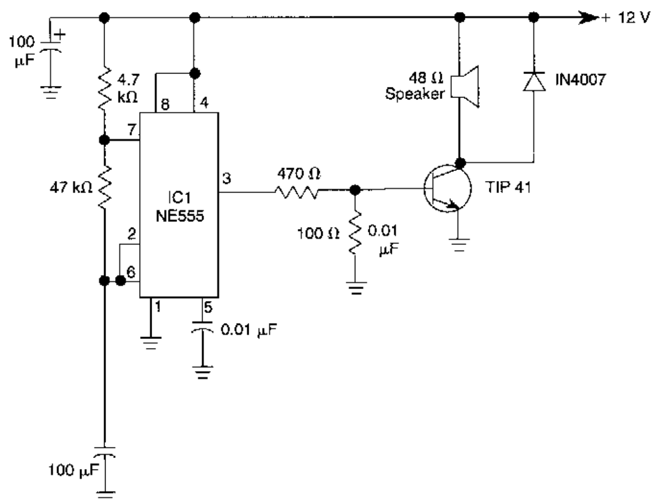
ELECTRONIC GONG



WILLIAM SHEETS ***FIG. 93-9***

The electronic gong is comprised of an oscillator (built around half of a 74COON quad 2-input NAND gate), an active twin-T filter (built around a TL081), and will drive an audio amplifier IC such as an LM386N. Pulses from astable multivibrator IC1 cause the twin-tee active filter U2 to ring, producing a damped sinusoidal output. C1 varies rate and C2-C3 vary gong frequency. Adjust R1 for best "tone" sound.

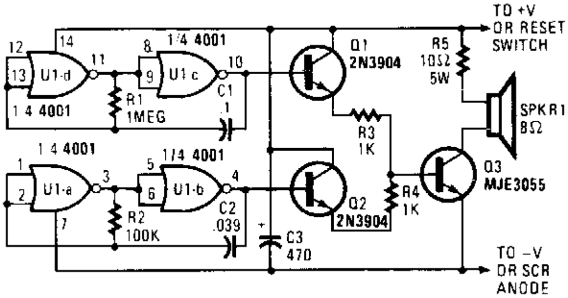
ALARM TONE GENERATOR



WILLIAM SHEETS ***FIG. 93-10***

In this alarm tone generator, a TIP41 transistor is used as a speaker driver. R1, R2, and C1 determines the frequency which is 1400 Hz with the values shown.

DUAL-TONE SOUNDER



POPULAR ELECTRONICS

FIG. 93-11

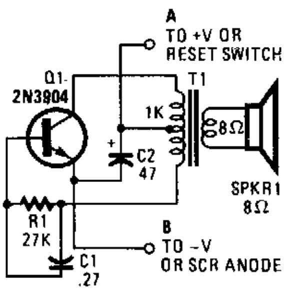
An outside horn-type speaker works best with the circuit. However, such devices require a great deal of power, so this sounder should only be used in alarm circuits where at least a 6-A SCR is used as the sounder driver.

A single CMOS 4001 quad 2-input NOR gate, two 2N3904 general-purpose npn transistors, and a single MJE3055 power transistor combine to generate a two-tone output. Gates U1-a and U1-b are configured as a simple feedback oscillator with R2 and C2 setting the oscillator's frequency. With the values shown, the circuit oscillates at about 500 Hz.

Gates U1-c and U1-d are connected in a similar oscillator circuit, but they operate at a much lower frequency. The oscillator frequencies (and thus the tones that they produce) can be altered by increasing or decreasing the values of R_1 and C_1 for the low-frequency oscillator and R_2 and C_2 for the high-frequency oscillator. Decreasing the values of those components will increase the frequency; increasing their values will decrease the frequency.

The two oscillator outputs are connected to separate amplifiers (configured as emitter followers), whose outputs are used to drive a single power transistor (Q3, an MJE3055). A 10-Ω, 5-W resistor, R5, is used to limit the current through the speaker and Q3 to a safe level. To boost the sound level, R5 can be replaced with another speaker.

LOW-LEVEL SOUNDER



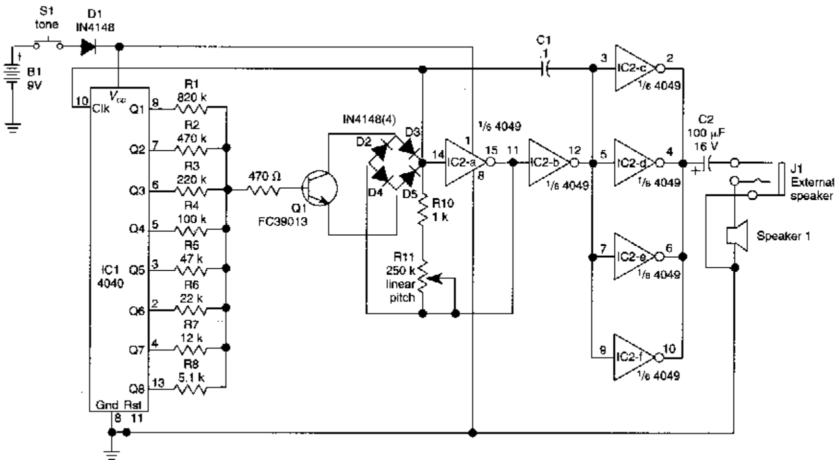
POPULAR ELECTRONICS

FIG. 93-12

This is a simple low-level noise maker that's ideally suited to certain alarm applications. When the sounder is located in another part of the building, the sound level is loud enough to be heard, but is not loud enough to warn off an intruder. A single 2N3904 npn transistor is connected in a Hartley audio oscillator, with a 1 kΩ to 8-Ω transistor-output transformer doing double duty.

The circuit produces a single-frequency tone that can be varied in frequency by changing the value of either or both R_1 and C_1. Increasing the value of either component will lower the output frequency and decreasing their values will raise the frequency. Don't go below 4.7 kΩ for R1 because you could easily destroy Q1.

SOUND-EFFECTS GENERATOR



1989 R-E EXPERIMENTERS HANDBOOK ***FIG. 93-13***

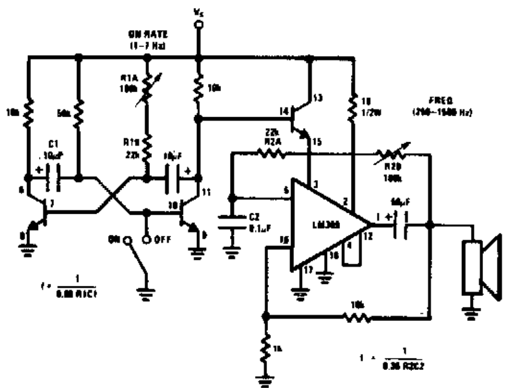
The circuit consists of four parts: a binary counter, a D/A converter, a VCO, and an audio output amplifier. The speed at which the counter counts depends on the frequency of the output of the VCO, which in turn is determined by the output of the counter. That feedback loop gives this circuit its characteristic output.

The initial frequency of oscillation is determined by potentiometer R11. The VCO first oscillates at a relatively low frequency, and it gradually picks up speed as the control voltage supplied by the D/A converter increases.

The D/A converter is simply the group of resistors R1 through R8. When none of IC1's outputs is active, little current will flow into the base of Q1, so the VCO's control voltage will be low. As more and more counter outputs become active, base current increases, and so does the VCO's frequency of oscillation.

The VCO itself is composed of IC2-a, IC2-b, and Q1; the timing network is D1 through D4, C1, R10, and R11. The diode bridge functions basically as a voltage-controlled resistor. The buffer amplifier is made up of the four remaining gates from IC2, all wired in parallel. The volume is sufficient for experimental purposes, but you might want to add an amplifier, speaker, or both.

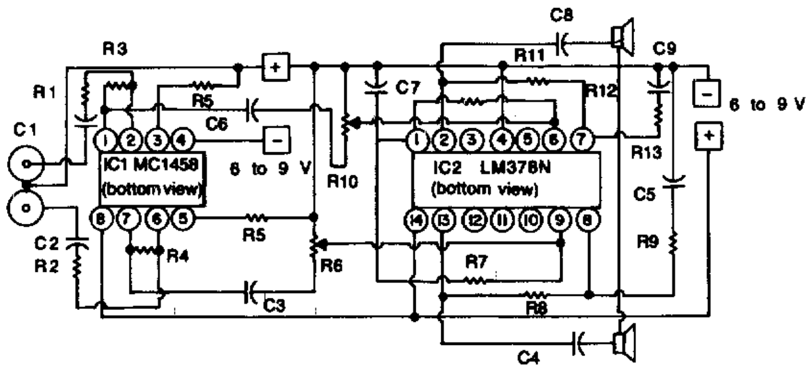
SIREN



An LM380 audio IC is configured as a feedback audio oscillator. A transistor astable modulates this oscillator at a low frequency, which produces a siren tone.

NATIONAL SEMICONDUCTOR ***FIG. 93-14***

ALTERNATE TONE ALARM



IC1	MC1458 dual op amp		
IC2	dual 4-watt amplifier	C1,C2	.1 μF
R1,R2	10 kΩ	C3,C6	.1 μF
R3,R4	1 MΩ	C4,C8	200 μF, 15 V
R5, R14	2.2 kΩ	C7	60 μF, 10 V
R10,R6	500 kΩ potentiometers	C9,C5	6 μF
R11,R7	100 kΩ	Spks	8 μF, 12-inch
R8, R12		Batteries (2 required or bipolar supply)	
R13,R9	4.7 kΩ		

1989 R-E EXPERIMENTERS HANDBOOK

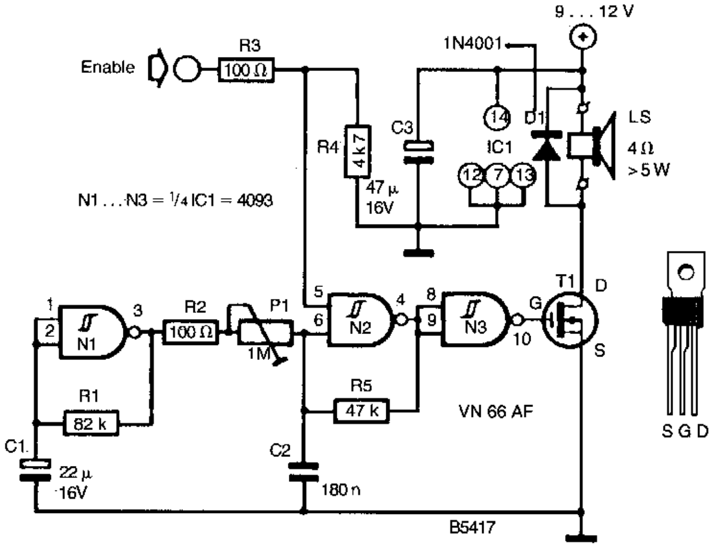
FIG. 93-15

A two-tone generator that is alternately switched ON provides a high/low output as might be heard from a traffic vehicle like a police car or ambulance.

IC1, CD4011, quad 2-input NAND gate is a two-tone oscillator in which each side, pins 1 through 7 and 8 through 13 set the tone frequencies. Changing the values of C_2 and C_1 determines the high/low tones. The outout frequencies are coupled to IC2, CD4011, of which one side (pins 1 through 6) acts as a buffer. The buffer is necessary to prevent loading on the outputs that would occur if one tried to go directly to the LM386 amplifier. The other side of IC2, pins 8 through 13, is a slow pulse oscillator of approximately 8 Hz per second. The output at pin 10 is connected to IC4 as a clock.

IC4, CD4027, is a dual J-K master-slave flip-flop that is wired to perform as a toggle switch in which Q1 and 15, and Q1 (NOT) pin 14, go high and low alternately (flip-flop). The clock input from IC2 pin 10 is connected to pin 13 of IC4, and the outputs at pins 15 and 14 changes the flip/flop state with each positive pulse transition. The CD4027 functions in toggle mode when the set and reset inputs, pins 9 and 12, are held low or grounded. Also, J-K inputs, pins 10 and 11, must be held high or to the positive. The outputs Q1 and Q1 (NOT), pins 15 and 14 are connected to pins 13 and 1 respectively of IC1 that enables or disables. Thus, each tone oscillator is turned on and off alternately. IC3 is a straightforward low-voltage audio amplifier.

SIREN OSCILLATOR



303 CIRCUITS

FIG. 93-16

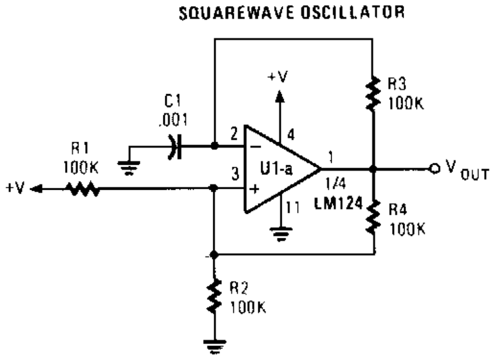
A CD4093 chip and a few components make up a siren oscillator, which drives power MOSFET T1. A 4-Ω speaker is driven directly from this device. The siren is enabled by a logic high applied to the ENABLE input.

94

Square-Wave Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SQUARE-WAVE OSCILLATOR

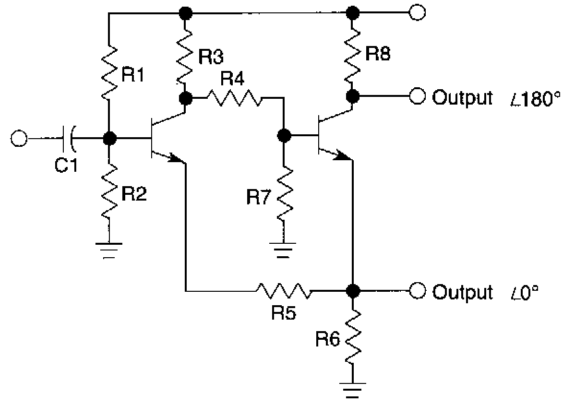


POPULAR ELECTRONICS ***FIG. 94-1***

An op amp with positive feedback generates a square wave. The period of the oscillator is determined by R3 and C1.

$$T = T_1 + T_2 \approx 0.69 \times 2\ (R_3C_1) \qquad T_1 = T_2$$

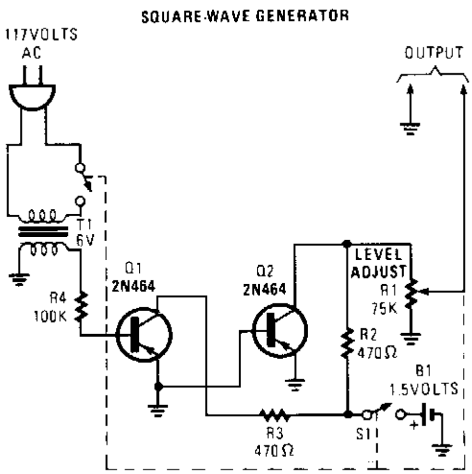
SCHMITT TRIGGER OR SINE-TO-SQUARE-WAVE CONVERTER



WILLIAM SHEETS ***FIG. 94-2***

This sine-wave triggered circuit produces two square-wave outputs that are 180° out of phase.

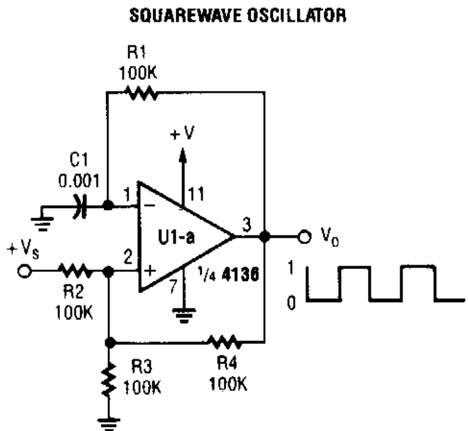
60-Hz SQUARE-WAVE GENERATOR



POPULAR ELECTRONICS ***FIG. 94-3***

This generator circuit uses an overdriven amplifier to produce a 60-Hz square wave from the 60-Hz ac line. The circuit can be used in line-operated applications as a clock source.

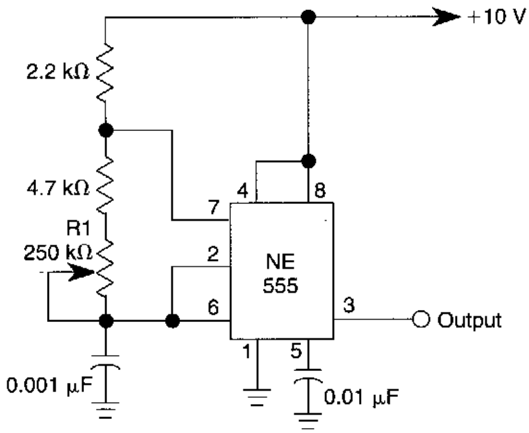
SQUARE-WAVE OSCILLATOR



POPULAR ELECTRONICS ***FIG. 94-4***

Positive feedback is via R3 and R4 and R1 and C1 determine period.

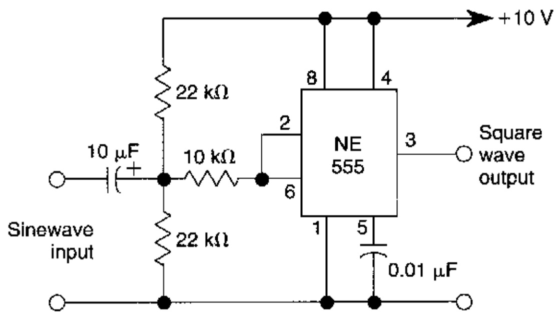
VARIABLE-FREQUENCY SQUARE-WAVE GENERATOR



WILLIAM SHEETS *FIG. 94-5*

This simple square-wave generator produces a variable frequency output of 2800 Hz to 80 kHz with the values shown. Frequency is adjusted with potentiometer R1.

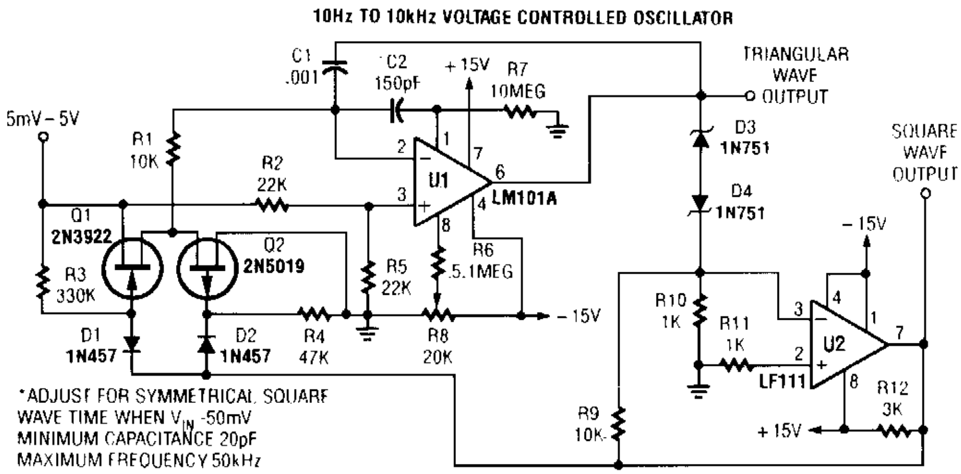
SCHMITT TRIGGER SINE-/SQUARE-WAVE GENERATOR



WILLIAM SHEETS *FIG. 94-6*

A sine wave input can produce a square wave output by this Schmitt trigger circuit based on a 555 IC.

10-Hz TO 10-kHz VCO WITH SQUARE- AND TRIANGLE-WAVE OUTPUTS



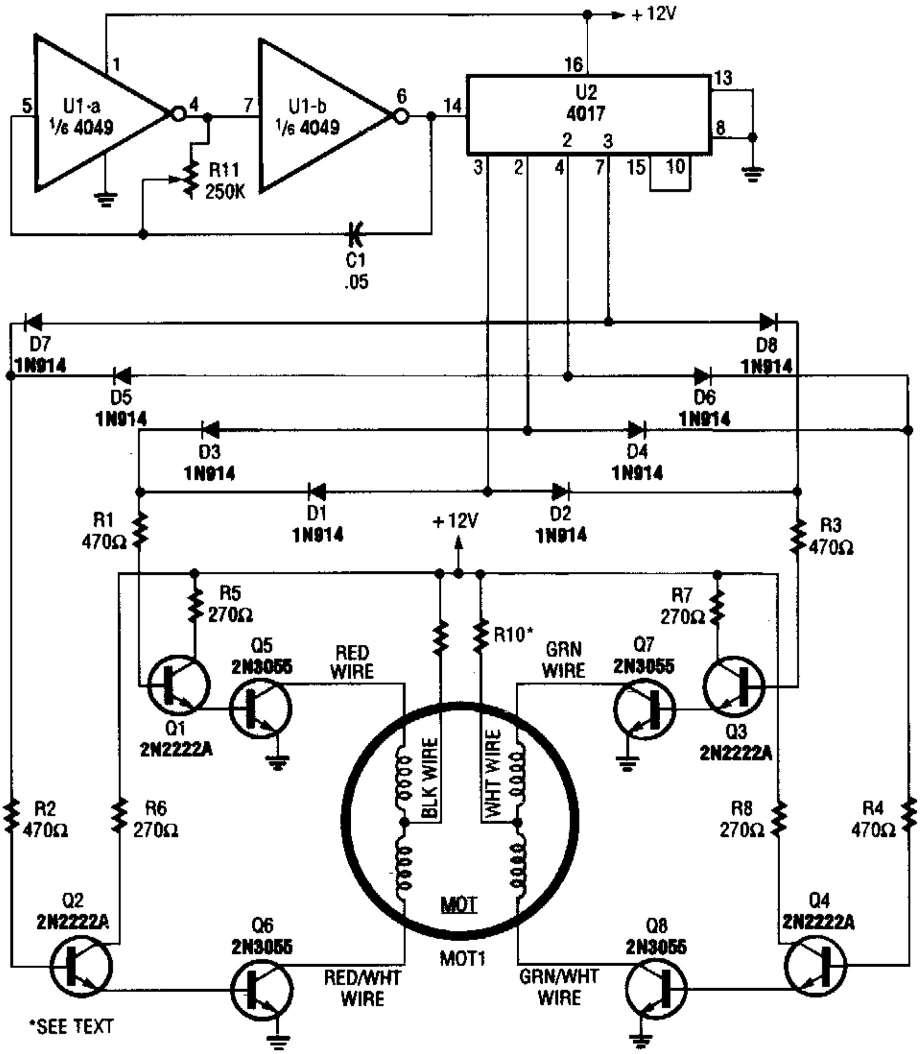
POPULAR ELECTRONICS *FIG. 94-7*

95

Stepper Motor Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

BIPOLAR STEPPER MOTOR DRIVE CIRCUIT

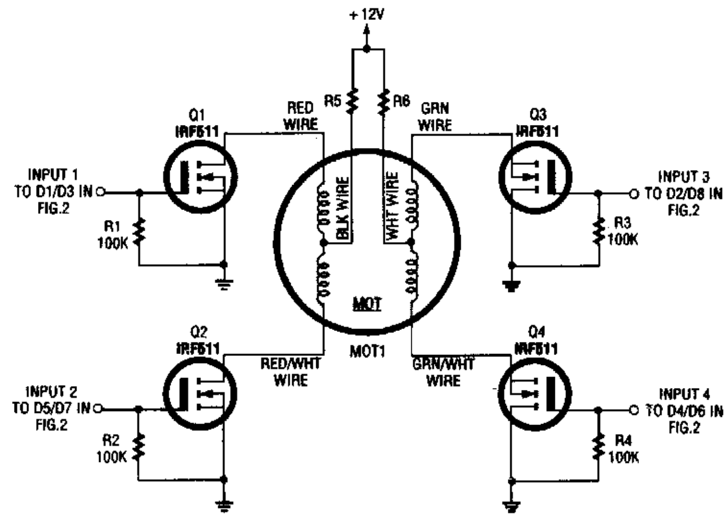


POPULAR ELECTRONICS

FIG. 95-1

A 4017 decade counter/divider driven from a low-frequency oscillator (U1-a and U1-b) is used to drive transistor switches to sequence the windings, as is needed. MOT1 is a 12-V stepper motor. R9 and R10 are selected for the motor's current rating. A 3.3-Hz signal from U1 will cause the motor to run at 1 rpm, a 33-Hz signal will result in 10 rpm, etc.

STEPPER MOTOR CIRCUIT WITH FET DRIVERS

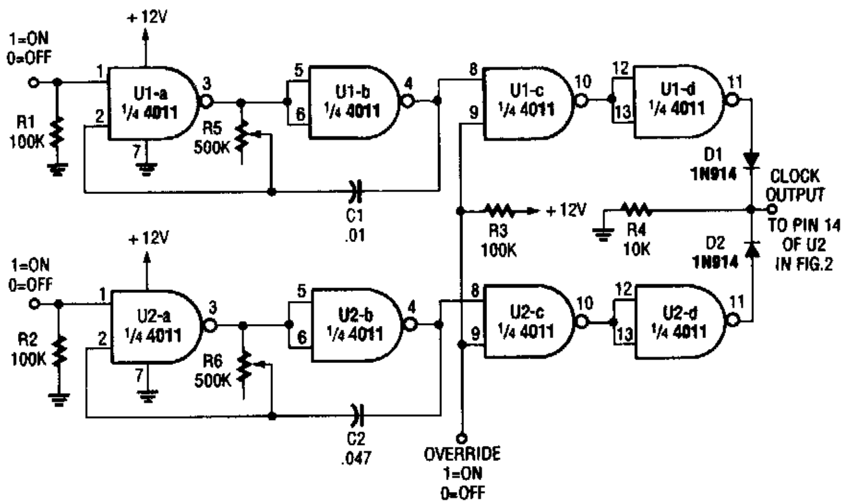


POPULAR ELECTRONICS

FIG. 95-2

This motor-driver circuit replaces the eight bipolar transistors of the previous circuit with four IFR511 power hexFET's (Q1 through Q4).

DUAL CLOCK CIRCUIT FOR STEPPER MOTORS



POPULAR ELECTRONICS

FIG. 95-3

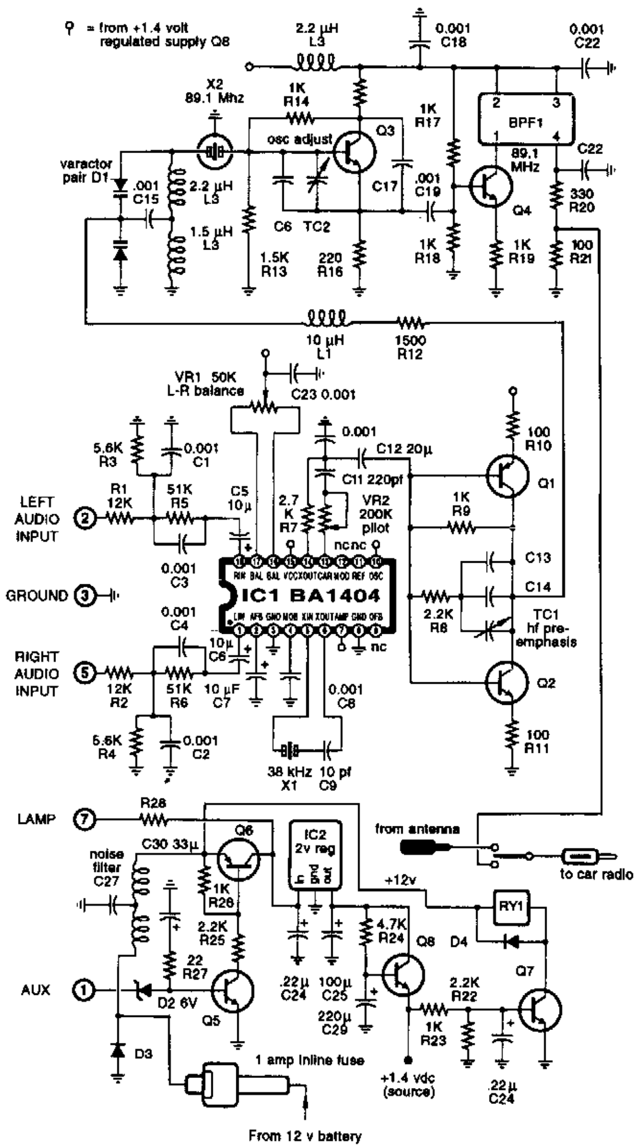
This oscillator can be used to drive a stepper motor circuit at two preset speeds with override to shut the motors off.

96

Stereo Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

FM STEREO TRANSMITTER

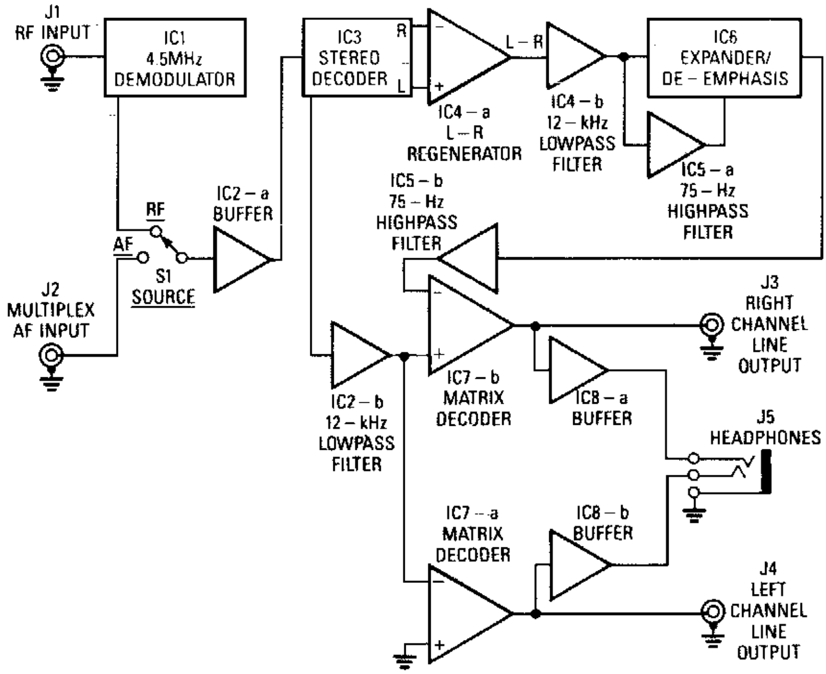


WARNING: Be sure to current limit the AUX input with an external 1K resistor!

RADIO ELECTRONICS **FIG. 96-1**

A BA1404 IC is used to generate a complete FM MPX signal. The chip contains all of the necessary circuitry. C1 and R3, and R4 and C4 provide pre-emphasis. The transmitter runs on a single AA cell. L3 is 3 turns of #20 wire on a 3/16" drill (for a form). L3 is 1/4" long. L4 is 4 turns #20 wire on 3/16" drill bit, spaced to 3/8". If monophonic operation is wanted, omit C5 and the 38-kHz oscillator components.

STEREO TV DECODER



1989 R-E EXPERIMENTERS HANDBOOK ***FIG. 96-2***

A block diagram of the stereo-TV decoder is shown in A. It shows the overall relationships between the separate sections of the circuit; B through E show the details of each subsection. The decoder section centers around IC1, a standard 4.5-MHz audio demodulator. The output of IC1 is routed to S1, which allows you to choose between the internally demodulated signal and an externally demodulated one. Buffer amplifier IC2-a then provides a low-impedance source to drive IC3, an LM1800 stereo demodulator.

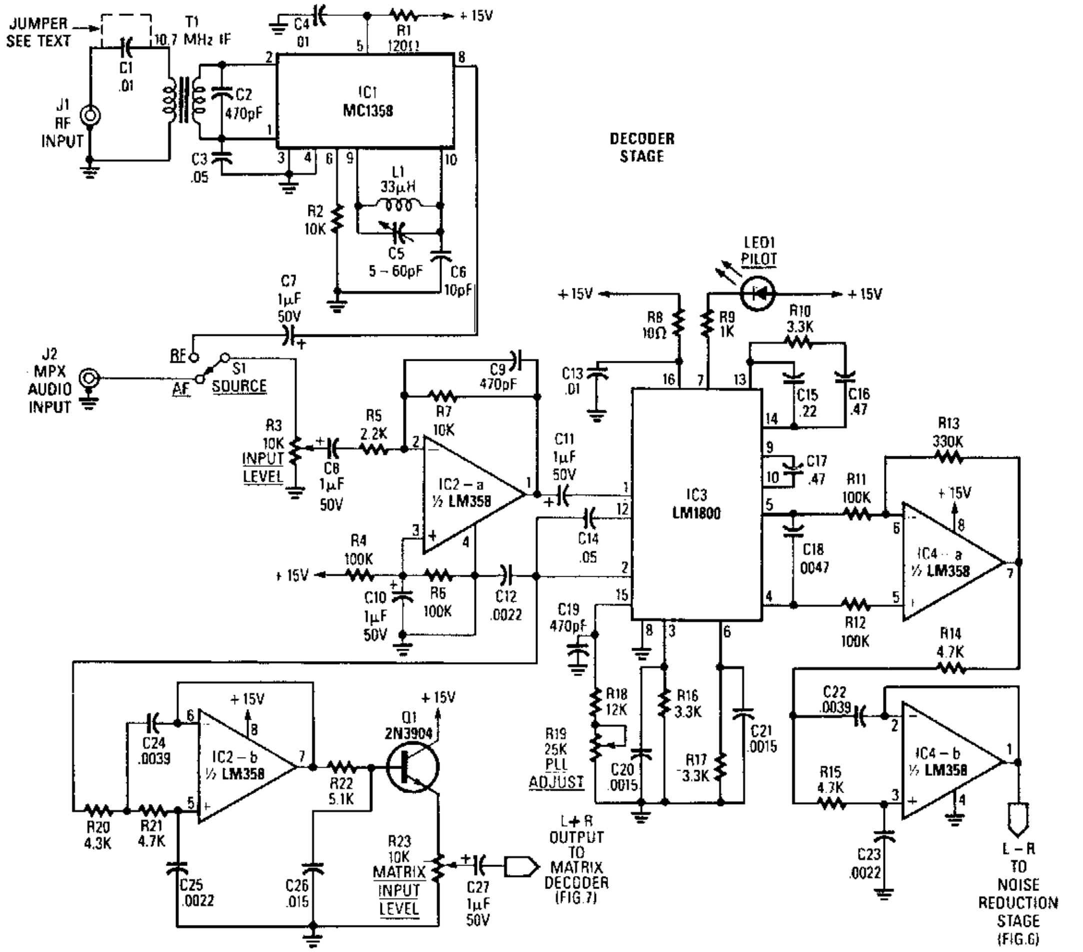
When IC3 is locked on a stereo signal, the outputs presented at pins 4 and 5 are discrete left- and right-channel signals, respectively. In order to provide noise reduction to the $L - R$ signal, you must recombine the discrete outputs into sum and difference signals. Op amp IC4-a is used to regenerate the $L - R$ signal. It is wired as a difference amplifier, wherein the inputs are summed together ($+L - R$). Capacitor C18 bridges the left- and right-channel outputs of the demodulator. Although it decreases high-frequency separation slightly, it also reduces high-frequency distortion.

The $L + R$ signal is taken from the LM1800 at pin 2, where it appears at the output of an internal buffer amplifier. The raw $L - R$ signal is applied to IC4-b, a 12-kHz lowpass filter. The $L + R$ signal is also fed through a 12-kHz low pass filter in order to keep the phase shift undergone by both signals equal.

Next, the $L - R$ signal is fed to Q2. It allows you to add a level control to the $L - R$ signal path; it provides a low source impedance for driving the following circuits, and it inverts the signal 180°. Inversion is necessary to compensate for the 180° inversion in the compander.

Next comes the expander stage. At the collector of Q2 is a 75-µs de-emphasis network (R27 and C29) that functions just like the network that is associated with Q1. Note that Q2 feeds both Q3 and

STEREO TV DECODER (*Cont.*)



THE DECODER STAGE converts the multiplexed audio signal into L + R and L − R signals.

IC5-a, a –12-dB per octave high-pass filter. The output of that filter drives the rectifier input of IC6, an NE570. The 75-Hz high-pass filter at the rectifier input helps to prevent hum, 60-Hz sych buzz, and other low-frequency noise in the $L - R$ signal from causing pumping or breathing.

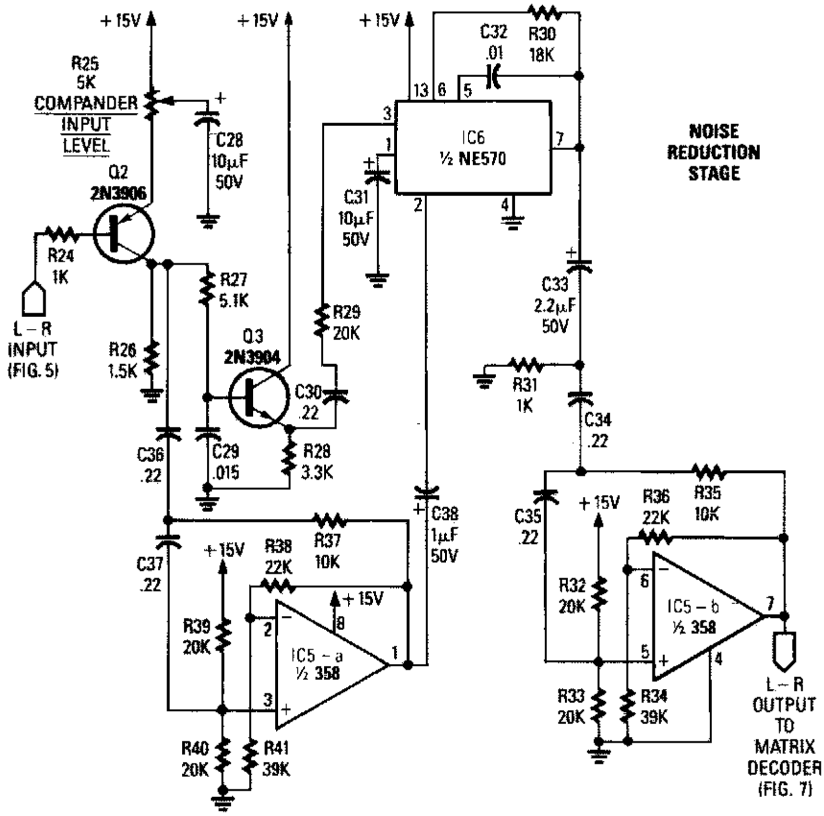
The NE570 contains an on-board op amp; its inverting input is available directly at pin 5 and via a 20-kΩ series resistor at pin 6. The 18-kΩ resistor (R30) combines with the internal resistor and C32 (0.01 μF) to form a first-order filter with a 390-μs time constant. Because the internal op amp operates in the inverting mode, the $-(L - R)$ signal is restored to the proper $(L - R)$ form.

The output of the expander drives another 75-Hz high-pass filter, but this one is a third-order type that provides –18 dB per octave rolloff. It is used to keep low-frequency noise from showing up at the output of the decoder. At this point, the $(L - R)$ signal has been restored, more or less, to the condition it was in before it was dBx companded at the transmitter.

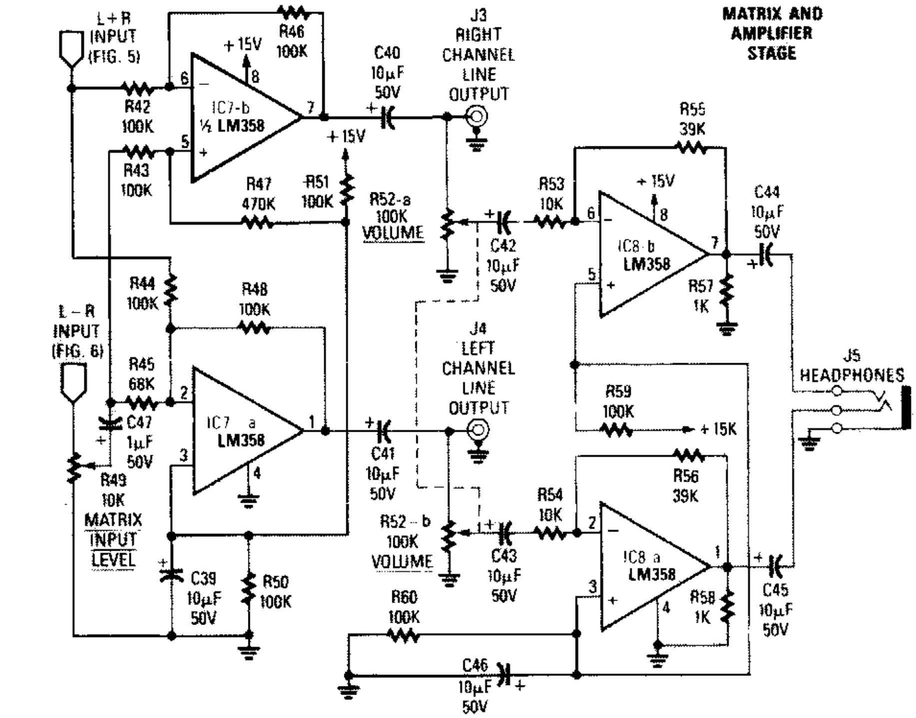
The $L + R$ signal from IC3 is fed to a 12-kHz low-pass filter, IC2-b, with a –12 dB per octave slope. The output of the high-pass filter is applied to a 75 µs de-emphasis network (R22 and C26). The $L + R$ audio signal is now restored properly. Q1 is wired as an emitter follower to provide a high load impedance for the de-emphasis network and a low source impedance for level control R23. Next, the $L + R$ signal is fed to the matrix decoder.

Op amps IC7-a and IC7-b are used to recover the individual channels. First, IC7-b is configured as unity-gain difference amplifier. The $(L + R)$ signal is applied to its inverting input, and the $(L - R)$ signal is applied to the noninverting input. Therefore, the output of IC7-b can be expressed as $-(L + R) + (L - R) = -L + L - R - R = -2R$. Similarly, IC7-a is configured as a mixing inverting amplifier. Here, however, both sum and difference signals are applied to the inverting input. So, the output of IC7-a is $(L + R) - (L - R) = -L - R - L + R = -2L$. Because both channels have been inverted, the stereo relationship is preserved.

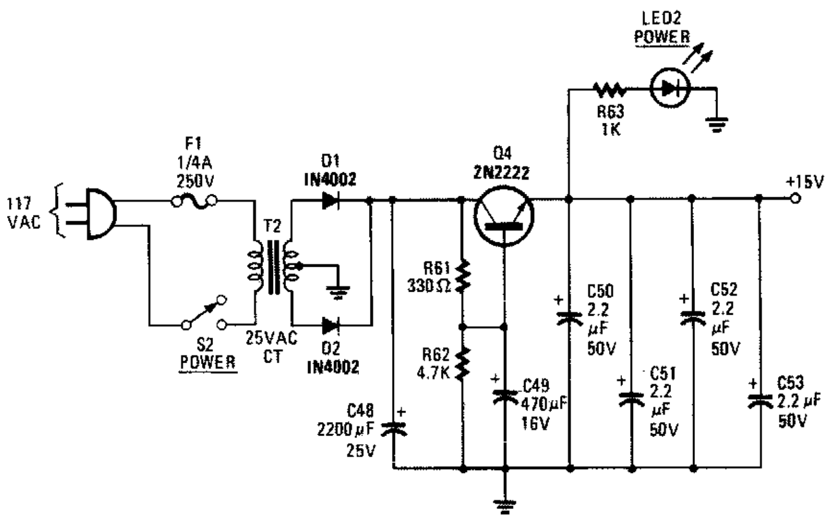
The two op amps in IC8 provide an additional stage of amplification to drive a pair of stereo headphones. If you don't plan to use your headphones, or if you are content to use only your stereo's headphone jack, all components to the right of line-output jacks J3 and J4 can be deleted.



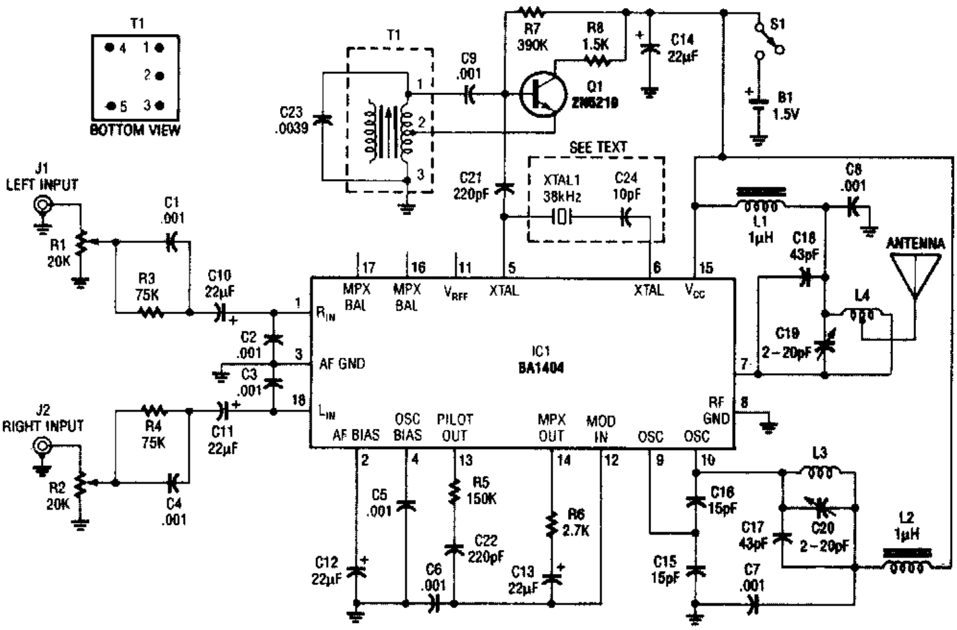
THE NOISE-REDUCTION STAGE de-compands the L – R signal, and emulates dbx-style processing. As described elsewhere in this article (see box), true dbx processing is not currently possible in a home-built circuit due to the inavailability of the dbx IC's.



THE MATRIX STAGE separates the L + R and L − R signals into the left- and right-channel components. Op-amp IC8 and associated components provide an optional headphone output. If you do not wish to drive a pair of headphones, or plan to use your amplifier's headphone jack for that purpose, all components to the right of jacks J3 and J4 can be deleted.



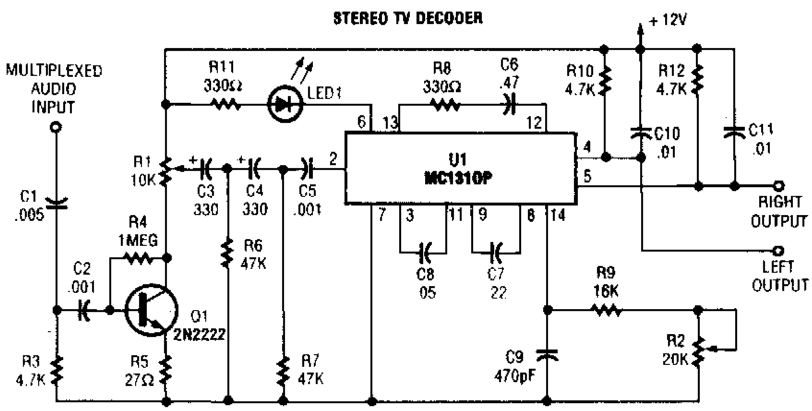
CRYSTAL-CONTROLLED FM STEREO TRANSMITTER



ELECTRONICS NOW *FIG. 96-3*

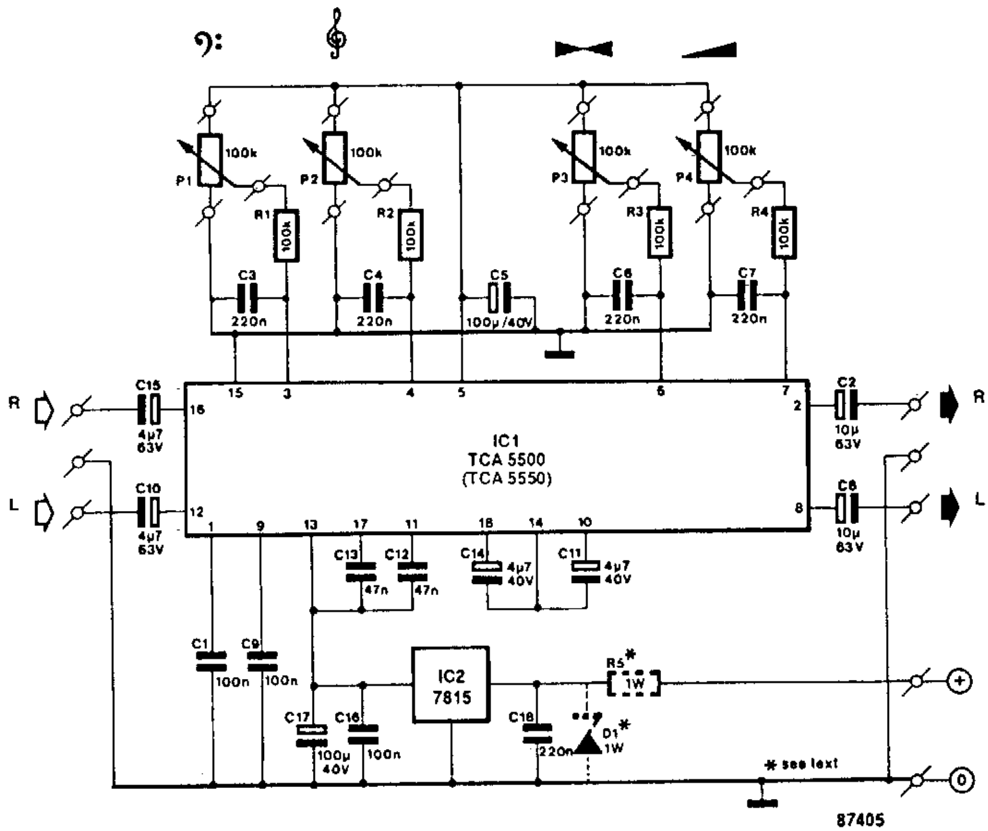
In this application, a BA1404 is used to generate an FM MPX baseband signal. This modulates a crystal oscillator (Q3) via a dual varactor series modulator. This transmitter can be to play CD audio on an existing FM auto radio.

STEREO TV DECODER



POPULAR ELECTRONICS *FIG. 96-4*

Q1 is an audio amplifier and U1 is used as a 31.5-kHz subcarrier, which is similar to 38-kHz FM MPX. Pilot frequency is 15.734 kHz.

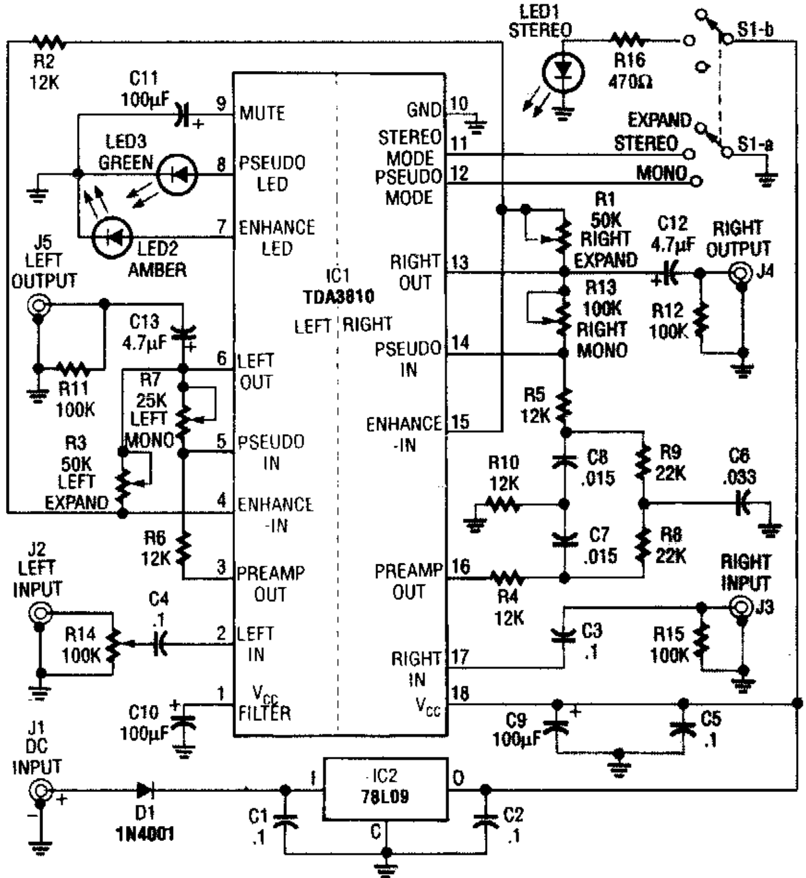


303 CIRCUITS

FIG. 96-5

A Motorola TCA5500 or TCA5550 can provide a stereo preamplifier system with tone controls. This circuit provides a gain of about 10X, a 14-dB tone-control range, a 75-dB volume control range, and it can operate from 8 to 18 Vdc. IC2 provides 15 V for IC1, and the input of IC2 can be supplied from the power amplifier's power supply (+) rail. D1 and R5 should be used if over 30 V input will be used.

AUDIO EXPANDER

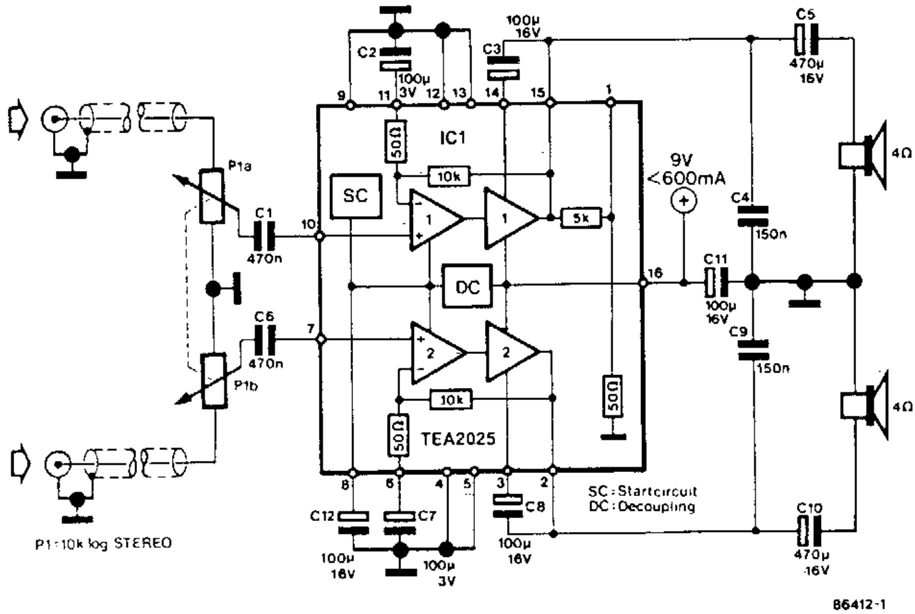


ELECTRONICS NOW

FIG. 96-6

This audio processor is based on the Signetics/Philips TDA3810N stereo, spatial, pseudo-stereo processor, IC. This processor uses a Philips TDA3810IC device, and it functions as an expander, pseudo stereo processor, and audio enhancer. Pseudo stereo is obtained by routing various frequencies to each channel via active filters.

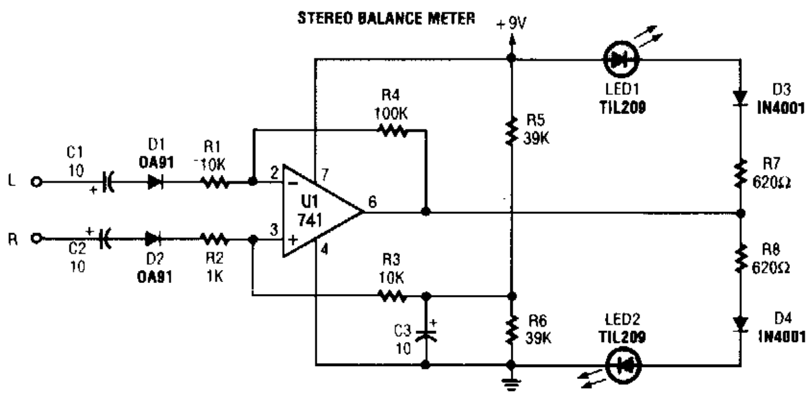
MINI STEREO AMPLIFIER



303 CIRCUITS ***FIG. 96-7***

Using a Thomson TEA2025, this stereo amplifier provides 1 W per channel into 4 Ω with a 9-V supply. Input sensitivity is 25 mV p-p for full output. Note that pins 4, 5, 12, and 13 of IC1 should be effectively grounded to a ground plane and heatsinked.

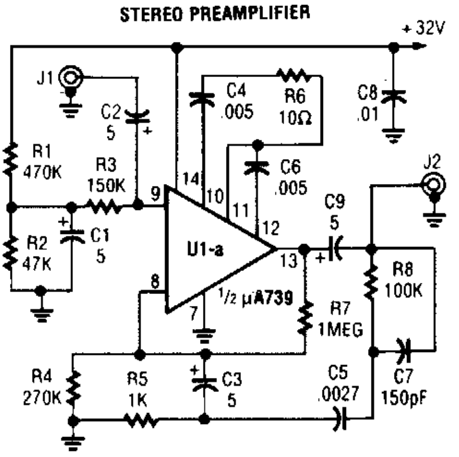
STEREO BALANCE METER



POPULAR ELECTRONICS ***FIG. 96-8***

When L & R signals are equal, no output is present from U1, and pin 6 is at a steady 4.5 V. Unbalanced audio causes the LEDs to vary in brightness, which causes a difference that corresponds to unbalance between channels.

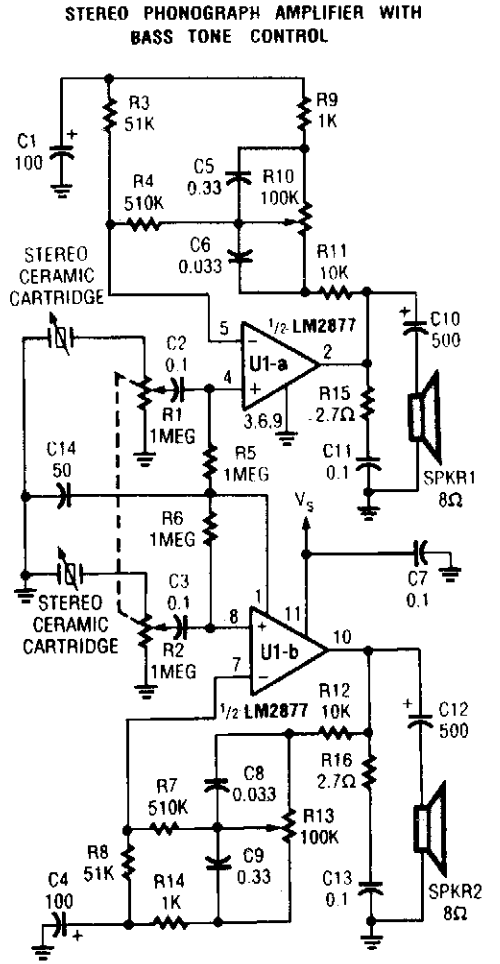
STEREO PREAMPLIFIER



POPULAR ELECTRONICS *FIG. 96-9*

A building block for audio work, the circuit can be used as a general-purpose preamp. Use two circuits for stereo applications.

STEREO PHONO AMPLIFIER WITH BASS TONE CONTROL



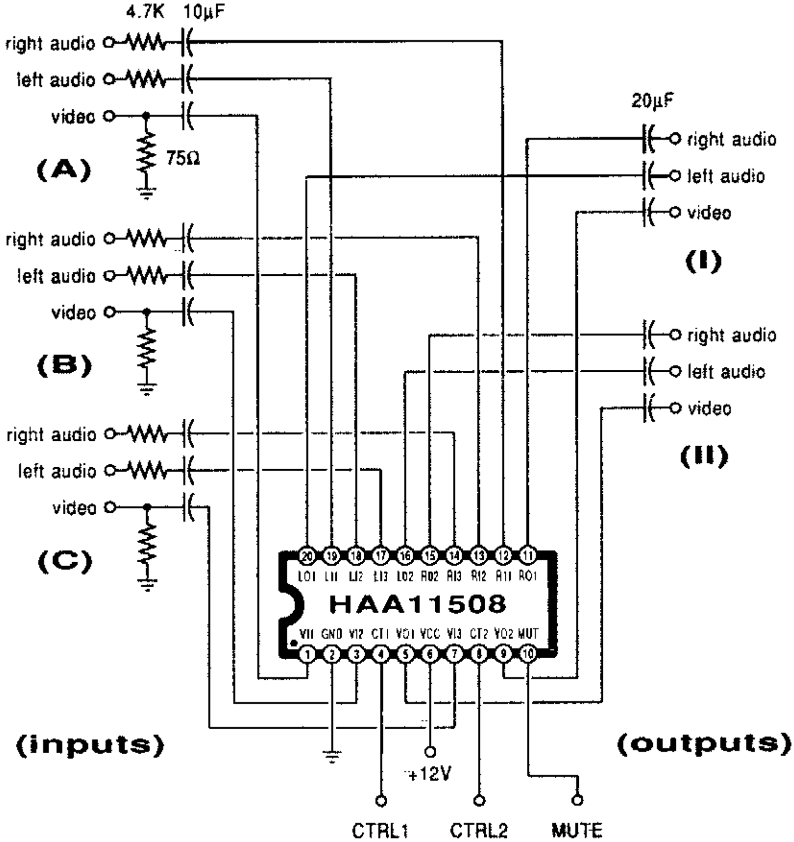
POPULAR ELECTRONICS *FIG. 96-10*

97

Switching Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE VIDEO/AUDIO SWITCHER

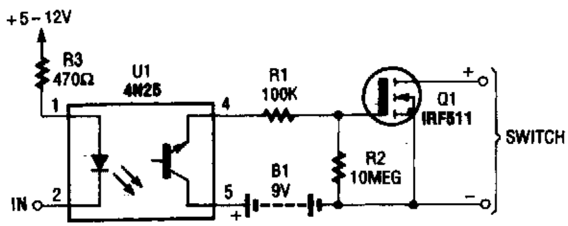


RADIO ELECTRONICS

FIG. 97-1

This channel selector selects video and stereo audio from any one of three different sources. The circuit should be constructed on a PC board with plenty of ground plane to minimize noise.

dc-CONTROLLED SWITCH USING OPTOISOLATOR

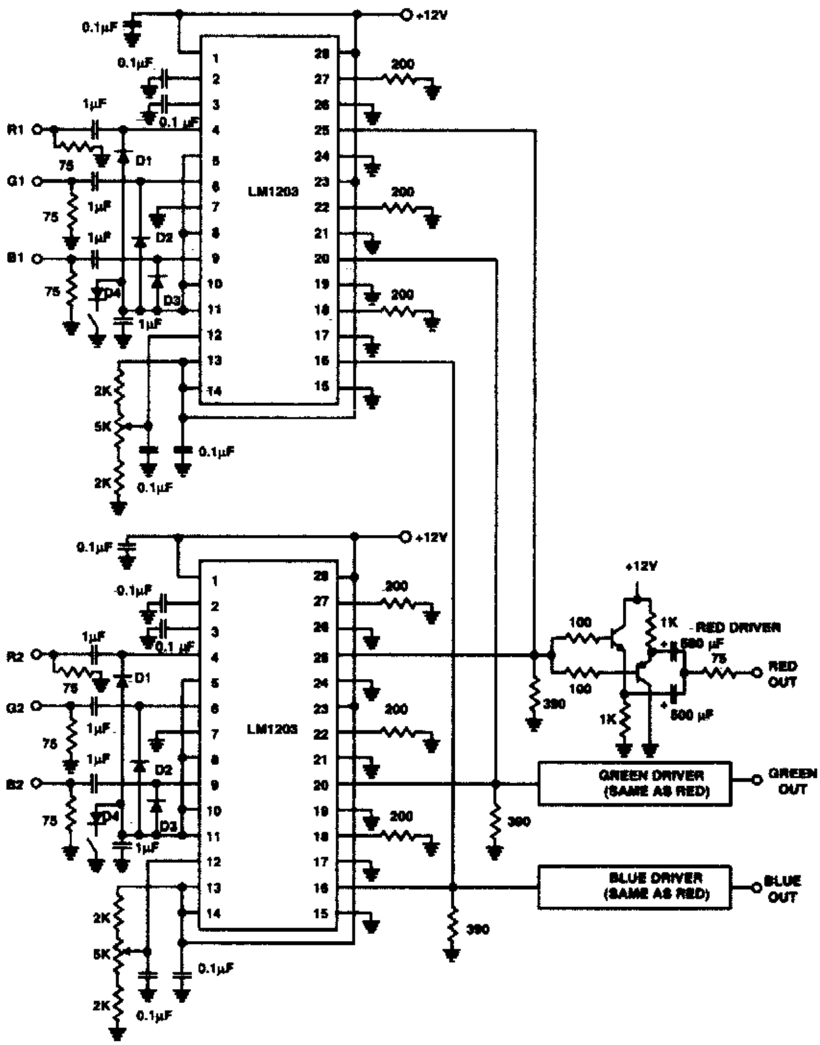


This dc-controlled switch uses an optoisolator/coupler, U1, to electrically isolate the input signal from the output-control device.

POPULAR ELECTRONICS

FIG. 97-2

WIDEBAND VIDEO SWITCH FOR RGB SIGNALS

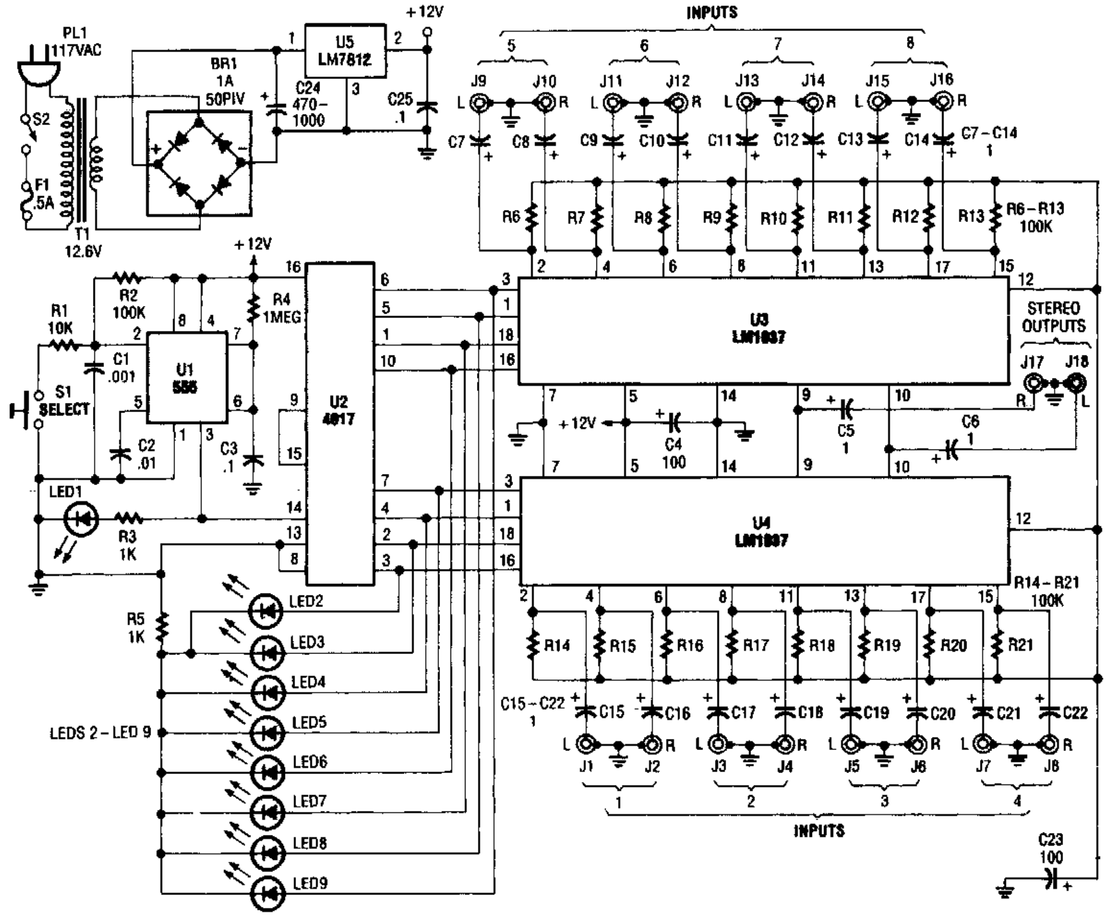


NATIONAL SEMICONDUCTOR

FIG. 97-3

The switch shown selects 1 to 2 inputs and uses a National LM1203. The slew rate is 4-V p-p into 390 Ω in 5 to 7 ns.

EIGHT-CHANNEL AUDIO SWITCHER



POPULAR ELECTRONICS ***FIG. 97-4***

This source is selected by pressing momentary-contact pushbutton switch S1. Switch S1 is connected to the trigger of a 555 oscillator/timer (U1) configured as a monostable multivibrator, which generates one short output pulse for each press of S1. That pulse turns on LED1 to give a visible indication that the 555 is working correctly. That pulse is also used to clock U2 (a 4017 CMOS divide-by-1-counter/divider).

Both LED1 and its associated current-limiting resistor R3 are optional and can be left out of the finished project without any affect on circuit operation. The 4017 advances by one clock pulse each time S1 is pressed, turning on its corresponding output. Pin 9 (corresponding to output 8) of U2 is directly connected to its own reset terminal at pin 15. This allows the counter to count from zero to seven, and then reset to zero on the eighth count.

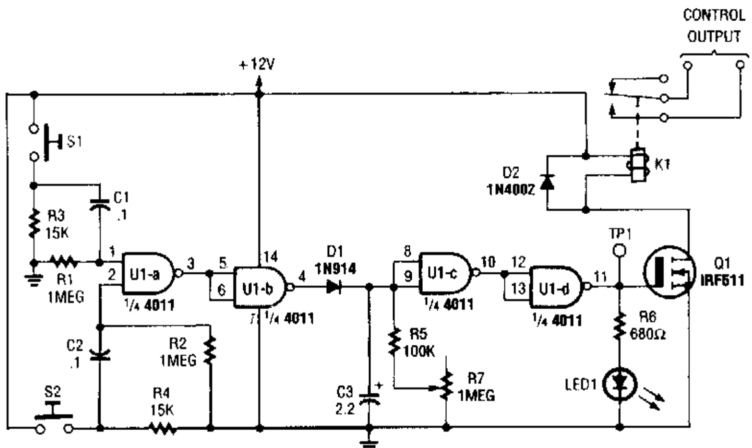
EIGHT-CHANNEL AUDIO SWITCHER (*Cont.*)

Pin 13, the enable input of U2, is tied to ground to allow the counter to operate. Outputs zero through seven are connected to eight indicator LEDs and the control pins of the two LM1037s (U3 and U4). When an output is selected, its LED lights and the corresponding control input on the LM1037 is brought high.

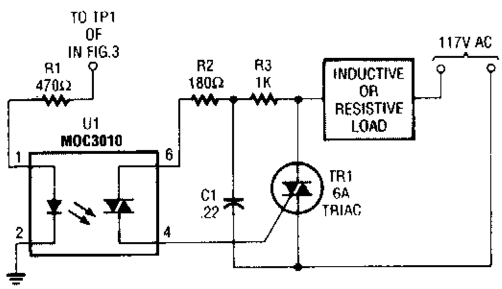
The LM1037 has extremely high-impedance inputs and low-impedance outputs, so interconnection between various types and brands of equipment should not be a problem. That, together with a wide-frequency response and low distortion, makes it ideal for use with good-quality, home-entertainment systems. The prototype of the audio switcher has a usable frequency response of from just a few hertz to over 100 kHz.

Power for the switcher is provided by a rather simple circuit. Because the switcher only draws between 20 and 30 mA, a simple circuit using the popular 7812 or 78L12 (a low-power version) voltage regulator works quite well.

ELECTRONIC SAFETY SWITCH



The electronic safety-control is built around a 4011 quad two-input NAND gate and an IRF511 hexFET.

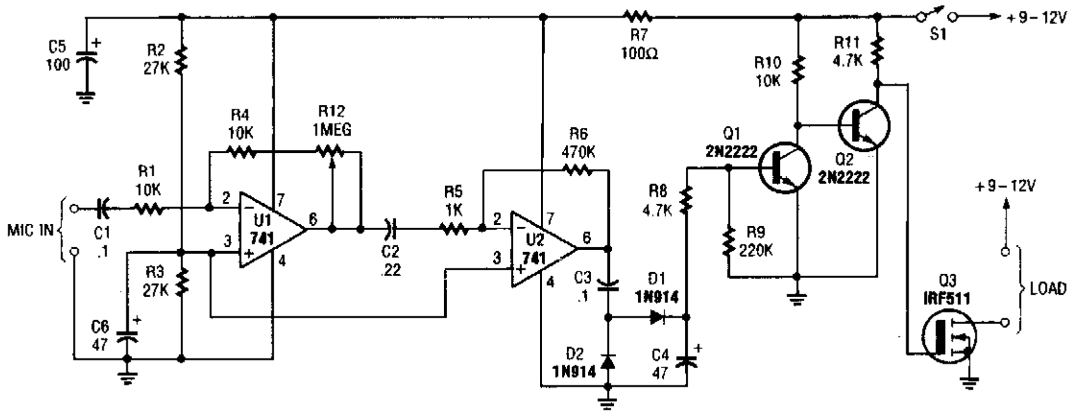


The relay-replacement circuit (shown here) can be used to operate inductive or resistive loads.

POPULAR ELECTRONICS **FIG. 97-5**

S1 and S2 must be depressed within 200 ms of each other to activate K1. The hold time is adjustable via R7. S1 and S2 overlap time can be changed by changing C1 and C2 or R1 and R2.

AUDIO-CONTROLLED SWITCH

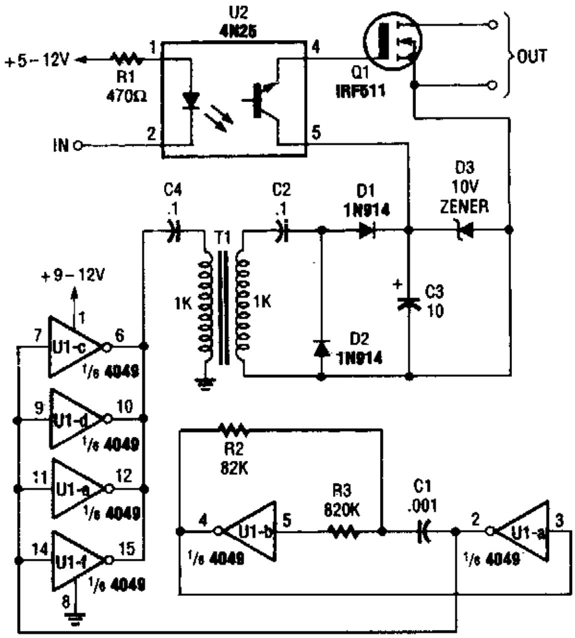


POPULAR ELECTRONICS

FIG. 97-6

This audio-controlled switch combines a pair of 741 op amps, two 2N2222 general-purpose transistors, a hexFET, and a few support components to a circuit that can be used to turn on a tape recorder, a transmitter, or just about anything that uses sound.

OSCILLATOR TRIGGERED SWITCH

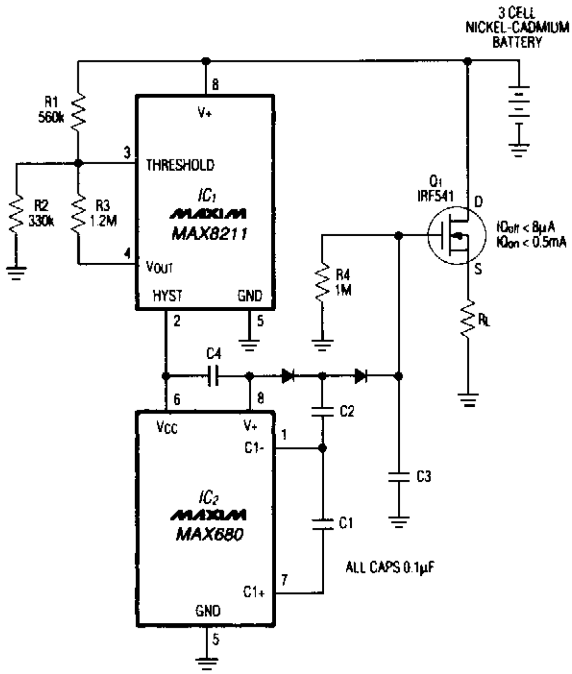


POPULAR ELECTRONICS

FIG. 97-7

An oscillator is used here to generate a 9-V bias to switch Q1. This removes the need for a battery as a bias source.

LOAD-DISCONNECT SWITCH

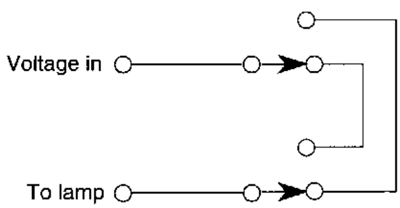


MAXIM ENGINEERING JOURNAL

FIG. 97-8

Deep discharge can damage a rechargeable battery. By disconnecting the battery from its load, this circuit halts battery discharge at a predetermined level of declining terminal voltage. Transistor Q1 acts as the switch. The overall circuit draws about 500 µA when the switch is closed and about 8 µA when the switch is open.

TYPICAL TWO-WAY SWITCH WIRING

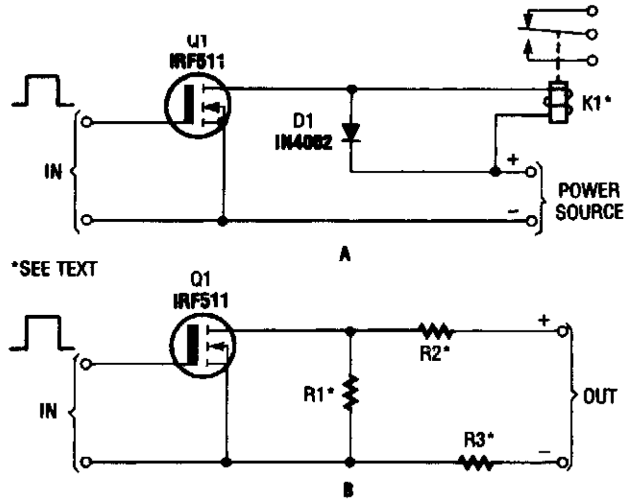


When the light is off, it can be turned on with either switch. When it's on, it can be turned off with either switch.

ELECTRONICS NOW

FIG. 97-9

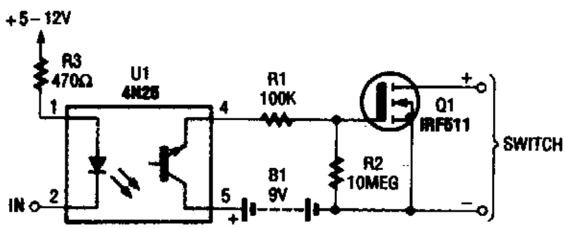
HEXFET SWITCH



POPULAR ELECTRONICS *FIG. 97-10*

The hexFET can switch dc power to relays (as shown in A), motors, lamps, and numerous other devices. That arrangement can even be used to switch resistors in and out of a circuit, as shown in B. R1, R2, and R3 represent resistive loads that can be switched in and out of the circuit.

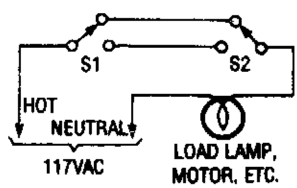
dc-CONTROLLED FET SWITCH



POPULAR ELECTRONICS *FIG. 97-11*

This dc-controlled switch uses an optoisolator/coupler, U1, to electrically isolate the input signal from the output-control device.

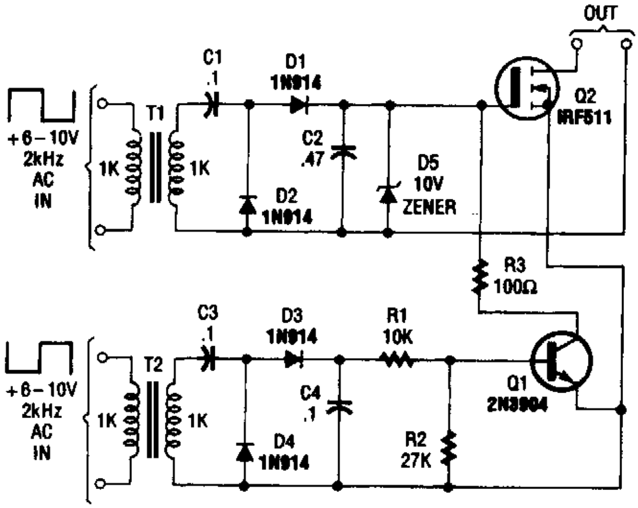
REMOTE TWO WAY ac SWITCH HOOKUP



POPULAR ELECTRONICS *FIG. 97-12*

This switching arrangement is the type of arrangement used in both domestic and industrial environments to allow a light or other ac-operated device to be controlled from more than one location.

DUAL-CONTROL HEXFET SWITCH



POPULAR ELECTRONICS *FIG. 97-13*

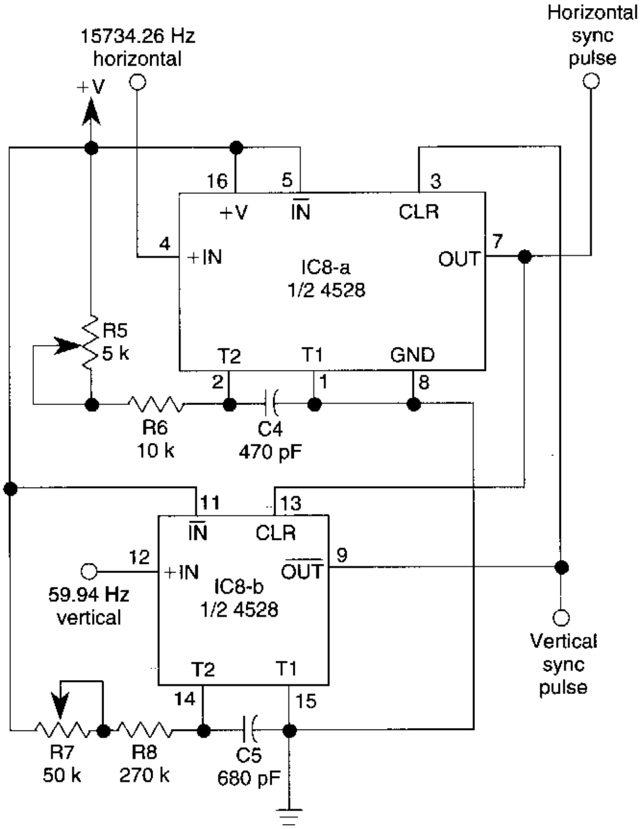
This dual-control switch uses two 6 to 10-Vac sources to trigger the circuit on and off; one source for each function.

98

Sync Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

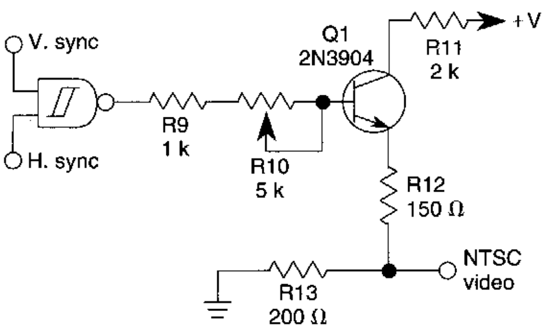
SYNC GATING CIRCUIT



RADIO ELECTRONICS ***FIG. 98-1***

This circuit guarantees that only one type of sync pulse is generated at a time. During vertical sync periods, horizontal sync is disabled.

SYNC COMBINER



This circuit combines H and V sync signals at TTL or CMOS levels and produces an NTSC video sync output.

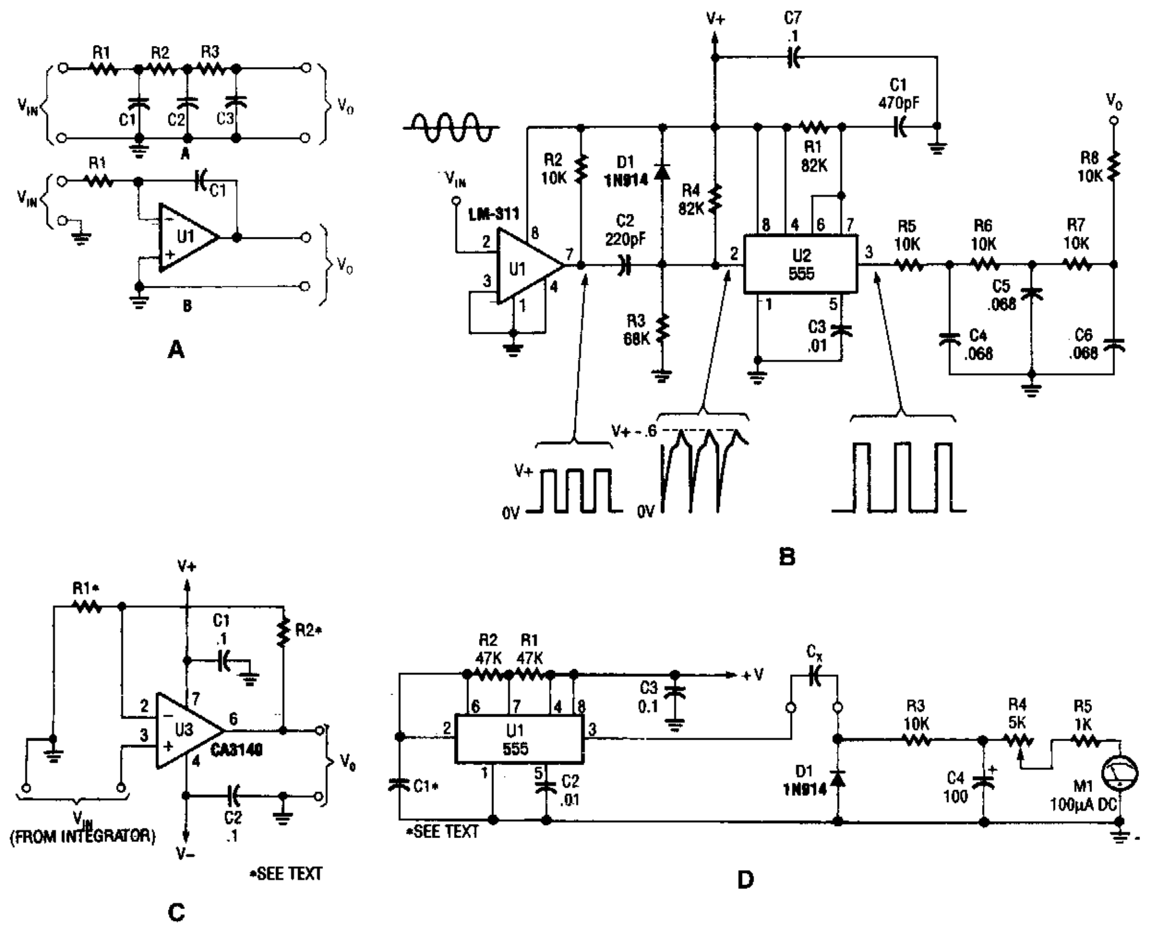
RADIO ELECTRONICS ***FIG. 98-2***

99

Tachometer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

ANALOG TACHOMETER CIRCUITS



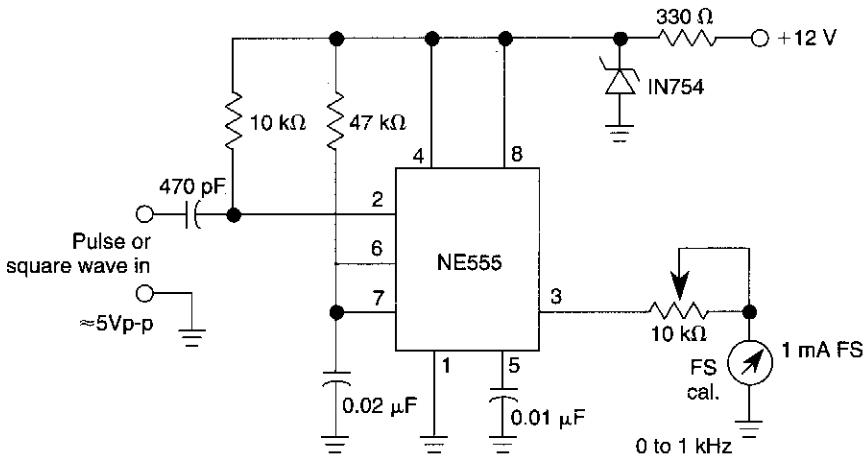
POPULAR ELECTRONICS **FIG. 99-1**

The four circuits shown are: a passive and active integrator, an analog tachometer, a scaling amplifier, and a capacitance meter.

In B, $T = 1.1\, R_1\, C_1$ (output pulse duration)

In C, $V_o = V_{in}\left(1 + \frac{R_2}{R_1}\right)$

ANALOG TACHOMETER CIRCUIT



WILLIAM SHEETS **_FIG. 99-2_**

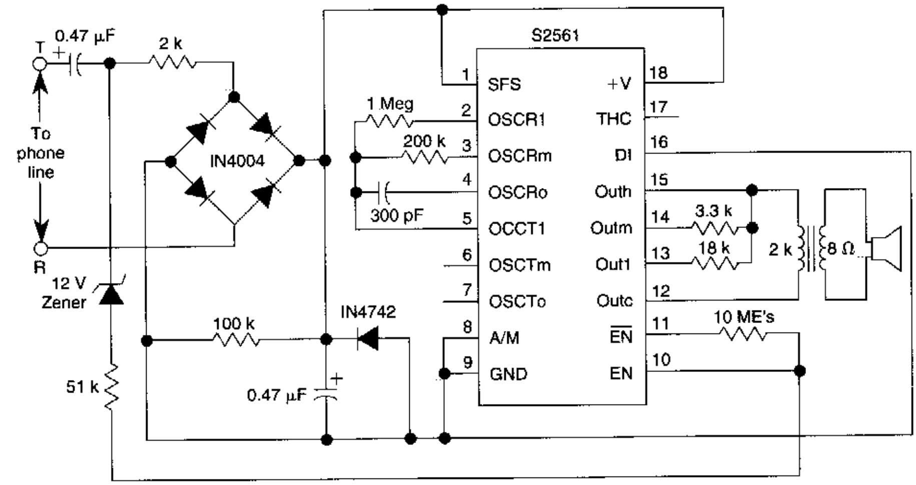
In this tachometer circuit a 555 is used as a pulse shaper. The dc value of the integrated pulse train is read by M1 which is calibrated to read frequency. With the values shown, the meter will read 0–1 kHz.

100

Telephone-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

TELEPHONE RINGER

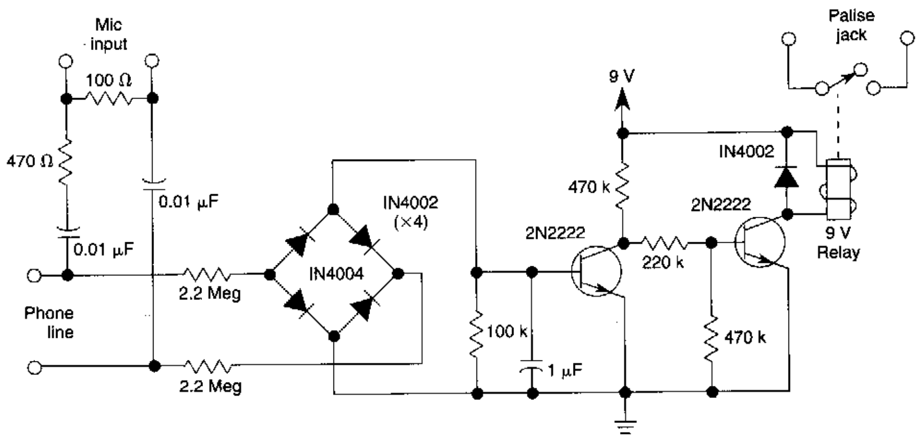


RADIO-ELECTRONICS

FIG. 100-1

Using an AMI chip P/N S2561, this telephone ringer can be powered directly off the telephone line. Audio output is about 50 mW when powered from a 10-V source.

AUTOMATIC TELEPHONE-CALL RECORDING CIRCUIT

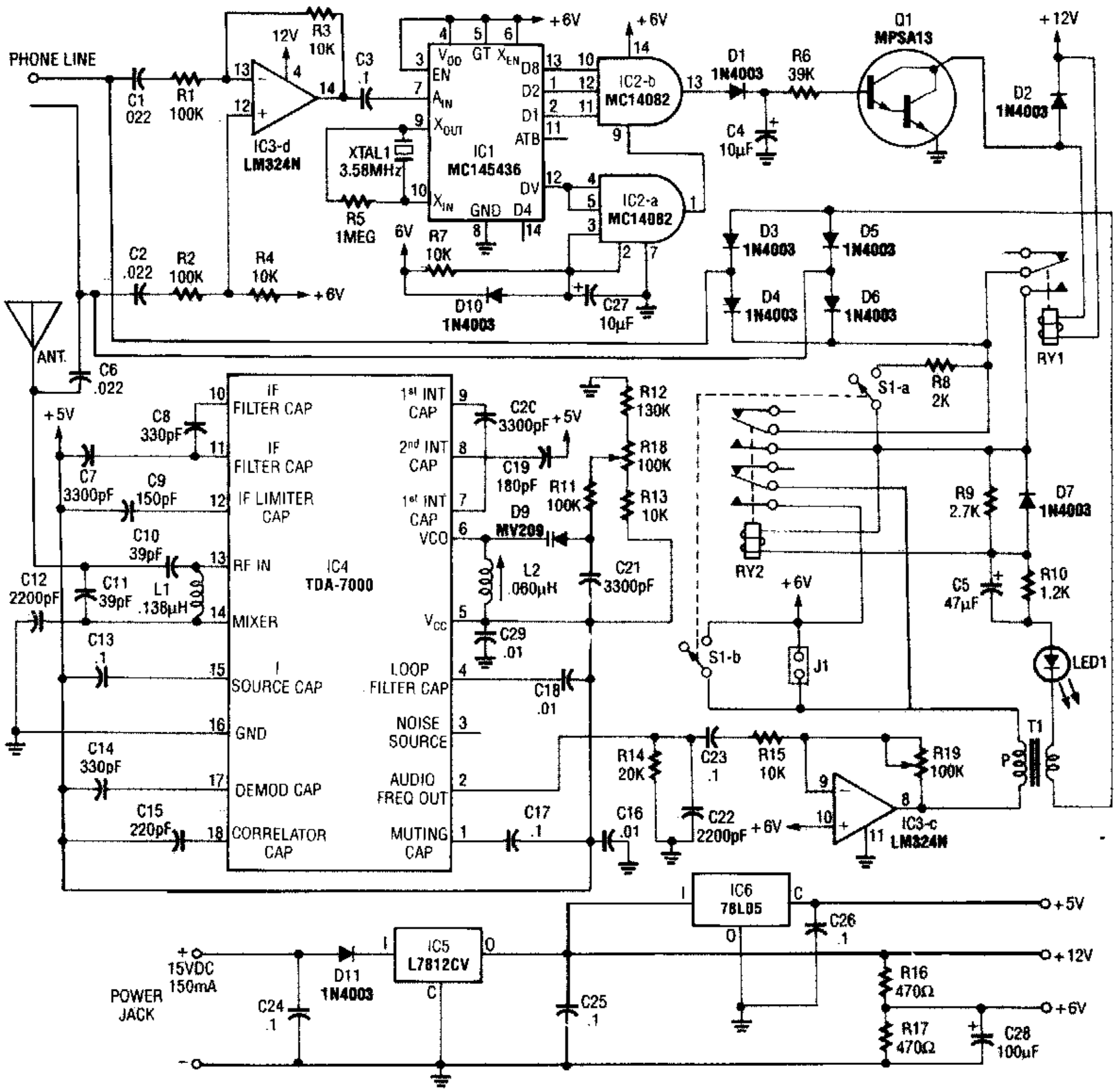


RADIO-ELECTRONICS

FIG. 100-2

The dc voltage present on a telephone line is usually around 45 to 50 V on-hook and 6 V off-hook. This circuit uses this drop in voltage to activate a relay. The relay controls a cassette tape recorder. Audio is taken off through a network to the microphone input of the cassette.

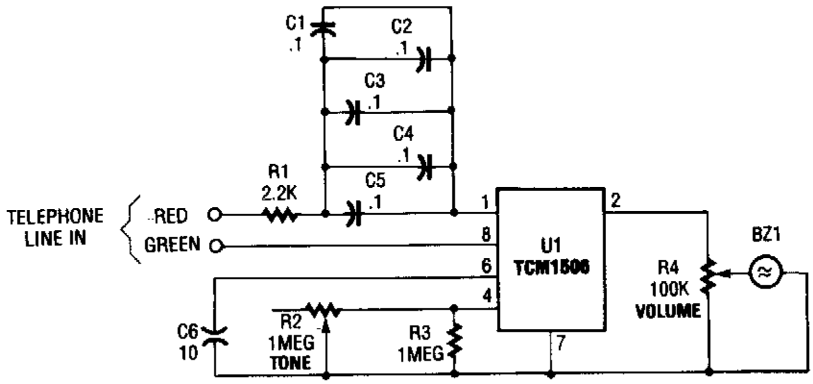
MUSIC ON HOLD



RADIO-ELECTRONICS *FIG. 100-3*

When an asterisk * is pressed on the touch-tone phone, IC1 a DTMF decoder, controls on-hold logic. Audio from the FM receiver IC4 is placed on the telephone line when a hold condition is present. RY2 is a DPDT 12-V relay. To place a caller on hold, press the asterisk button on the touch-tone phone and hang up the handset.

TELEPHONE RING CONVERTER



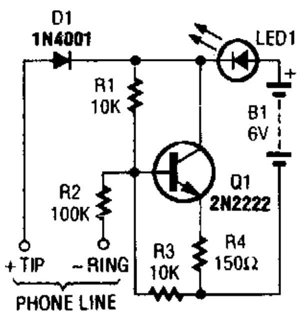
POPULAR ELECTRONICS *FIG. 100-4*

The circuit is based on the TCM1506 ring detector/driver integrated circuit. It is a monolithic IC specifically designed to replace the telephone's mechanical bell. The chip is powered and activated by the telephone-line ring, which can vary from 40 to 150 V rms at a frequency of from 15 to 68 Hz. No other source of power is required. Again, referring to the figure shown, C1 through C5 are placed in parallel to form a 0.5-µF capacitor that conducts the ac ring voltage to pin 1 of the TCM1506, but blocks any dc component. Of course, those capacitors can be replaced by a single 0.47- to 0.5-µF capacitor provided that it has at least a 400-WVdc rating. Resistor R1 is in series with the capacitor network and is used to dissipate power from any high-voltage transient that might appear across the line. The diluted ac voltage that reaches pin 1 on U1 powers the chip.

Capacitor C6 is used to prevent "bell tapping." That is an annoying ringing of the bell that occurs when a phone on the same line is used to dial an outgoing call. The capacitor prevents the short dial pulses from triggering the ring detector, but still allows the much longer ring signal to activate it.

Potentiometer R2 is used to vary the tone of the ring signal from below 100 Hz to over 15 kHz. Potentiometer R4 is the volume control; adjusting that potentiometer to its lowest resistance will mute the piezo element (BZ1). When a ring signal is present on the phone-line, it powers U1. The IC then generates a tone (with a frequency that is determined by R2 and an amplitude set by R4) that is reproduced by BZ1.

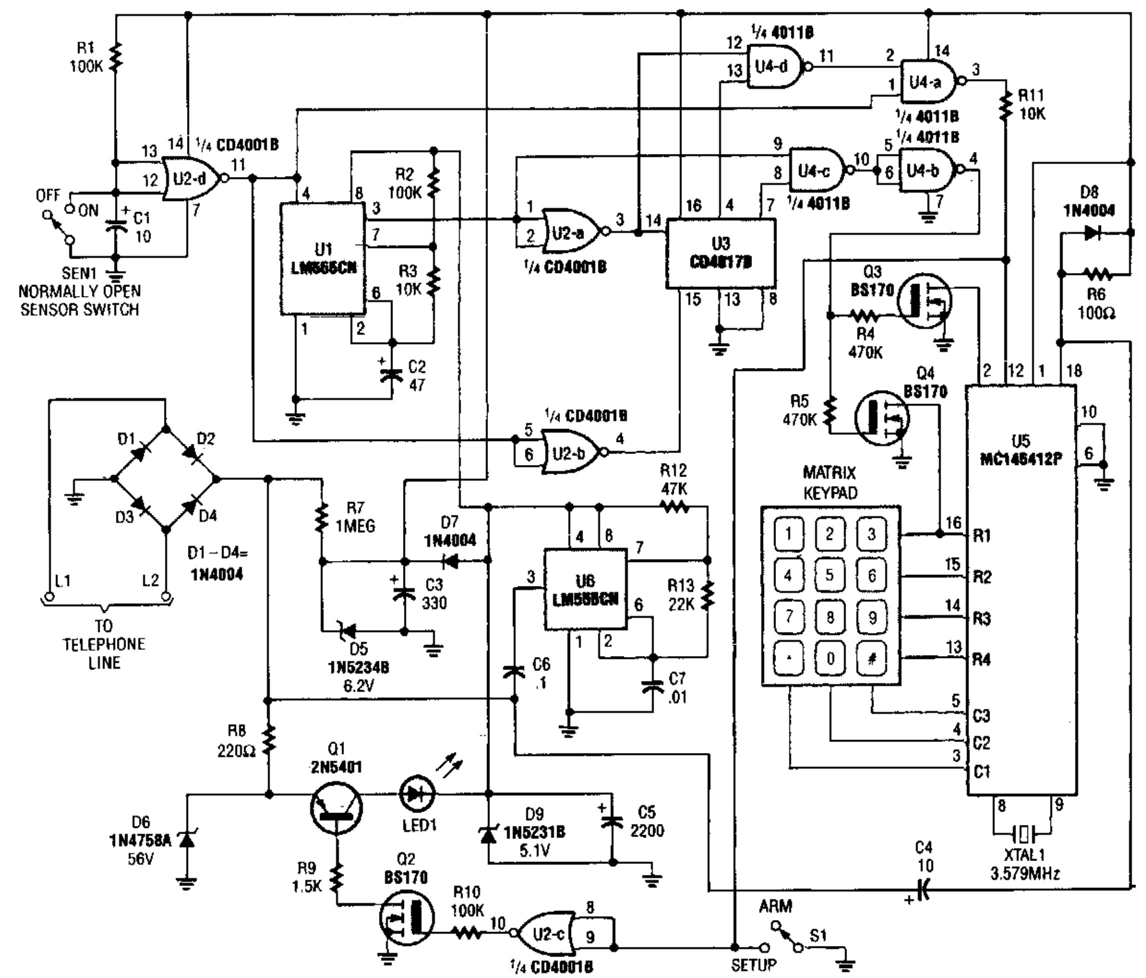
PHONE-IN-USE INDICATOR



This phone-in-use indicator also indicates the presence of a ring signal. Just the thing for the hearing impaired.

POPULAR ELECTRONICS *FIG. 100-5*

EMERGENCY TELEPHONE DIALER

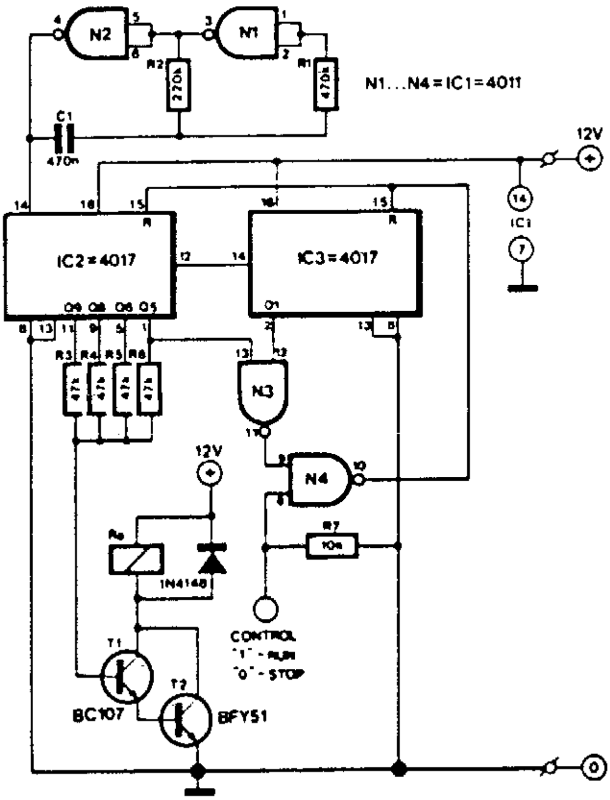


POPULAR ELECTRONICS ***FIG. 100-6***

This system will alert you or anyone chosen by automatically dialing a programmed phone number. This is accomplished by monitoring an open-loop or closed-loop sensor switch located in the protected area. When the sensor detects a problem (such as a break-in, fire, heating system failure, flood, etc.), Teleguard dials whatever telephone number has been programmed into its memory. When the phone is taken off the hook, Teleguard emits an unusual tone to alert the party on the receiving end that something is amiss.

The circuit is not hampered by busy signals when a call is placed; it automatically redials the number again and again (about once a minute) until it gets through. In addition, Teleguard can also automatically dial a number in the event of a medical emergency; for instance, where a mobility-impaired person is unable to dial the telephone. That can be accomplished by adding a "panic" switch to the circuit.

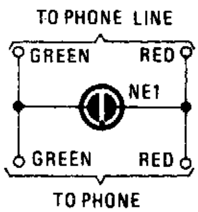
TELEPHONE BELL SIMULATOR



303 CIRCUITS ***FIG. 100-7***

This circuit is intended for use in a small private telephone installation. The ringing tone sequence is 400 ms on, 200 ms off, 400 ms on, 2 ms off. In the accompanying diagram, N1 and N2 form an oscillator that operates at a frequency of 5 Hz, which gives a period of 200 ms. The oscillator signal is fed to two decade scalers, which are connected in such a manner (by N3 and N4) that the input signal is divided by 15. The second input of N4 can be used to switch the divider on and off by logic levels. If this facility is not used, the two inputs of N4 should be interconnected.

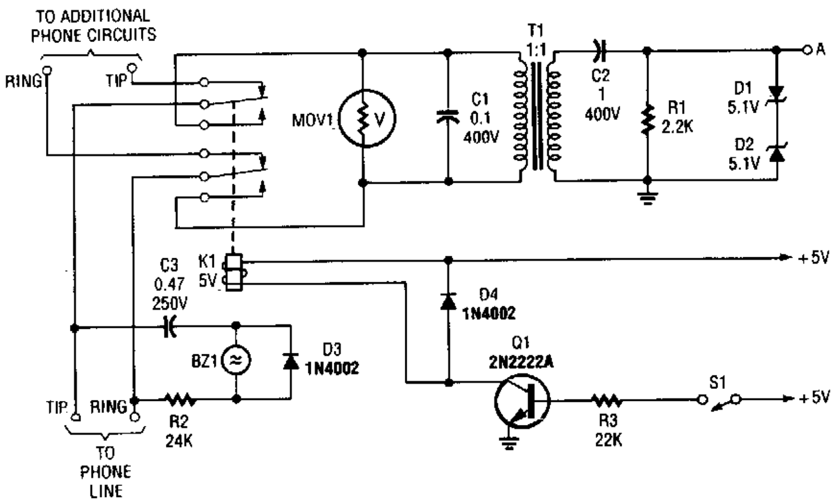
SIMPLE TELEPHONE RING INDICATOR



A neon lamp can easily be added to the phone line to act as a ring indicator. It's perfect for times when you can't hear the phone.

POPULAR ELECTRONICS ***FIG. 100-8***

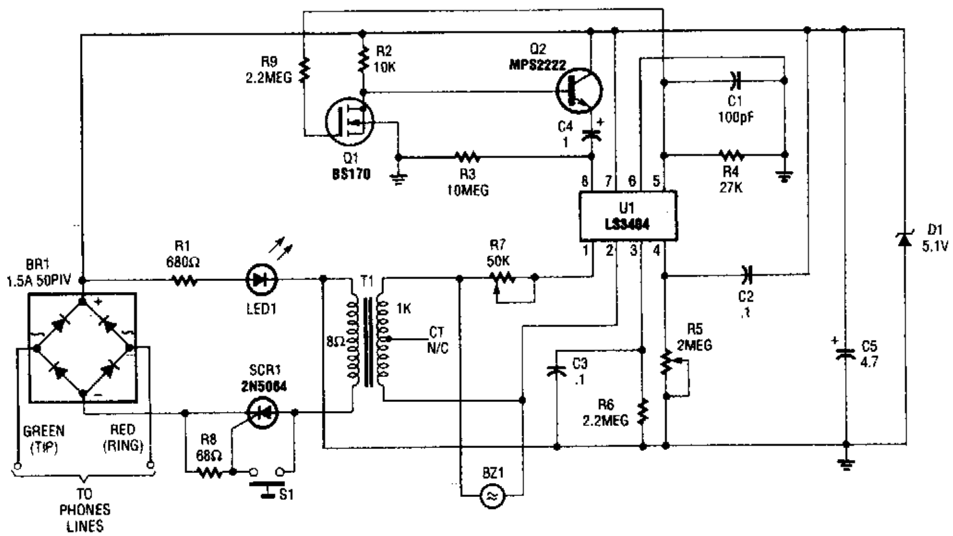
PHONE-LINE INTERFACE



POPULAR ELECTRONICS ***FIG. 100-9***

This circuit should be useful for interfacing phone projects to the telephone line. It has a ringer, can interrupt the wiring, and isolates project from the phone line.

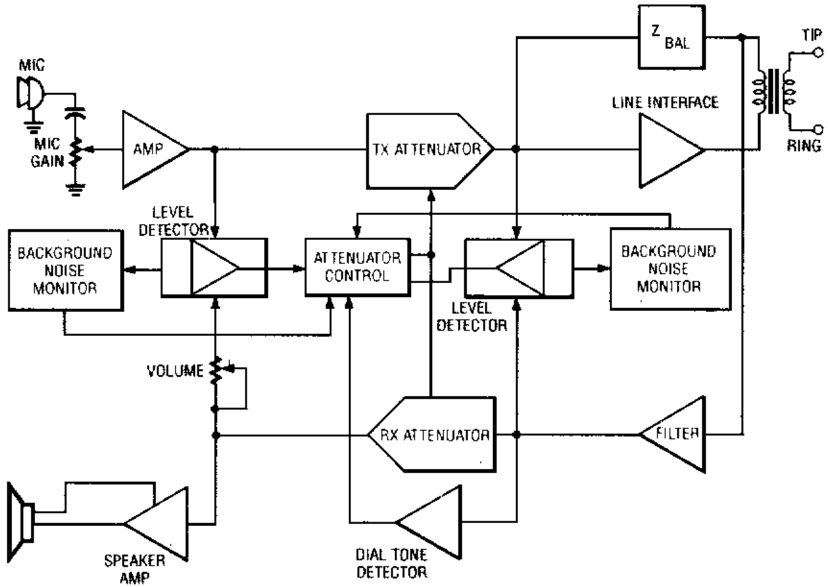
MUSIC-ON-HOLD BOX



POPULAR ELECTRONICS ***FIG. 100-10***

U1, an LS3404 melody chip is activated when "hold" S1 is pressed, which causes SCR1 to conduct and hold the telephone line via T1, R1, and LED1. The voltage across R1 and LED1 is used to activate the melody chip. Q1 and Q2 form a restart circuit to keep the melody chip going during hold.

SPEAKERPHONE ADAPTER



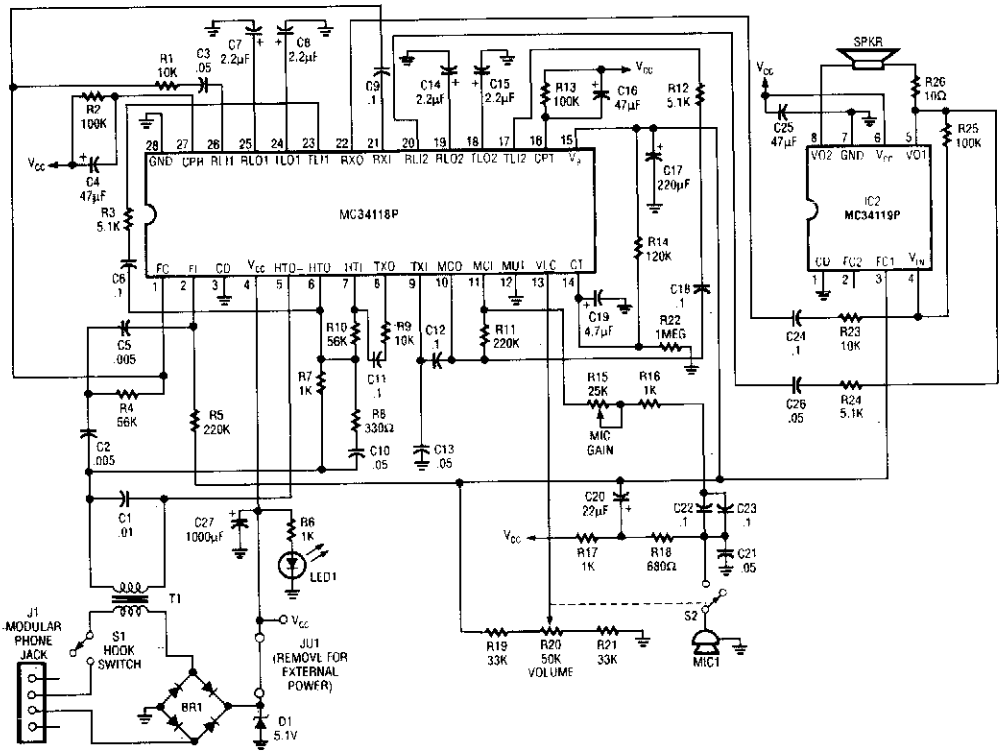
BLOCK DIAGRAM. **The talk path goes left to right on the upper half of the drawing, and the receive path goes from right to left.**

RADIO-ELECTRONICS

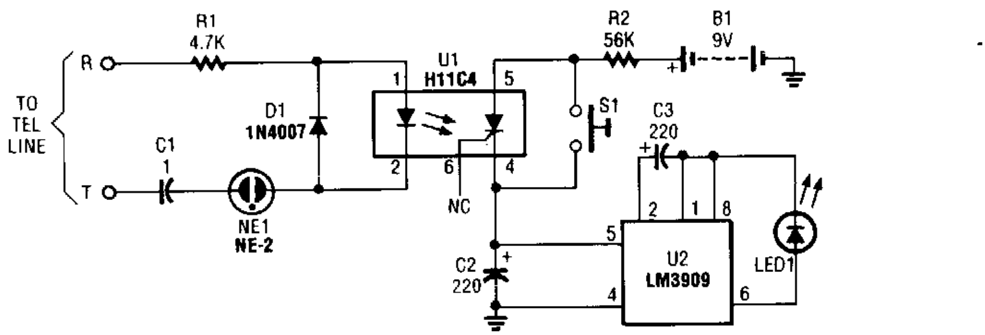
FIG. 100-11

Using a Motorola MC34118 speakerphone IC, this adapter can be used with a regular telephone to provide speaker capability. This device is powered from the phone line, but it can be powered via an external power supply if the line loop current is marginally low. An external phone is needed for ringing and dialing functions.

SPEAKERPHONE ADAPTER (*Cont.*)



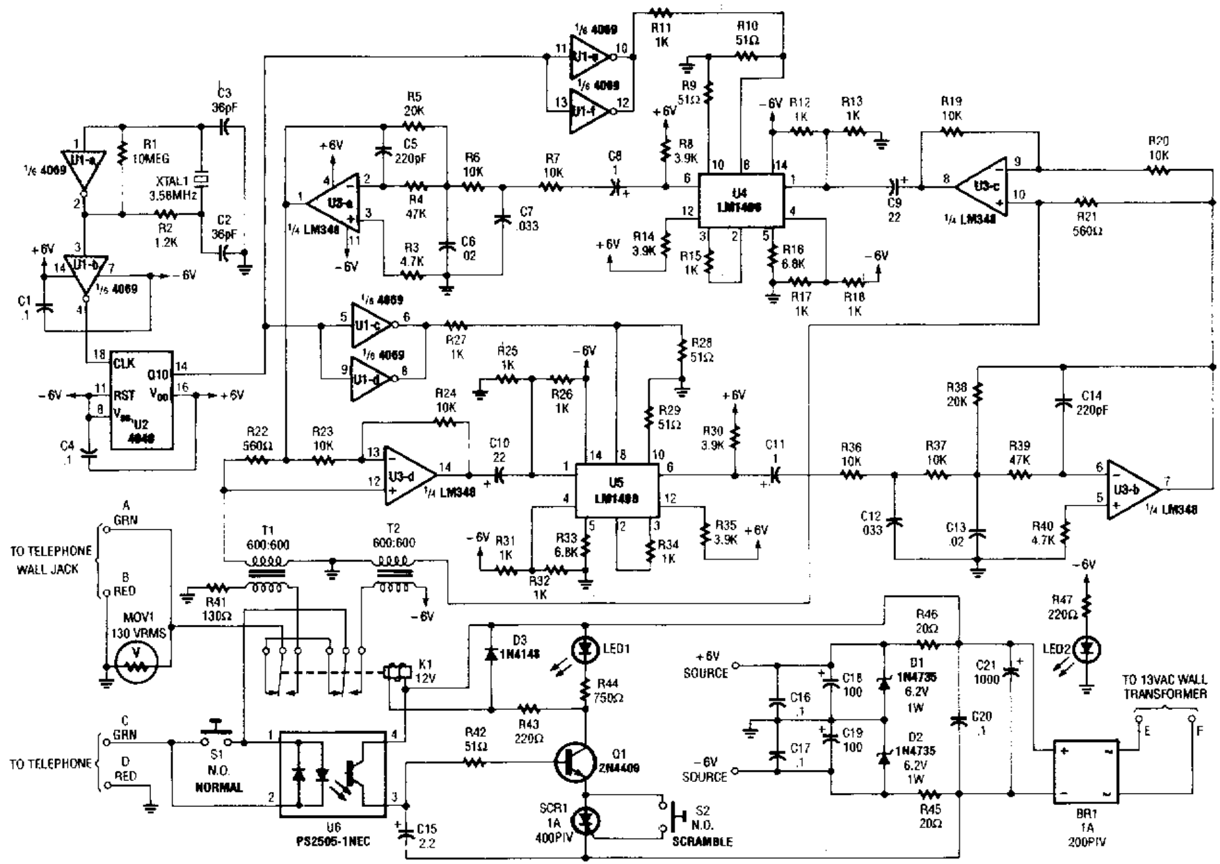
TELEPHONE VOICE-MAIL ALERT



1993 ELECTRONICS HOBBYIST HANDBOOK *FIG. 100-12*

The circuit is built around a couple of low-cost ICs: an H11C4 optoisolator/coupler with an SCR output (U1) and an LM3909 LED flasher (U2). It is connected to the phone line in the same manner as any extension phone. A ring signal on the telephone activates the optoisolator/SCR, and causes U2 to flash LED1. This flash signifies that a ring signal has been received.

TELEPHONE SCRAMBLER

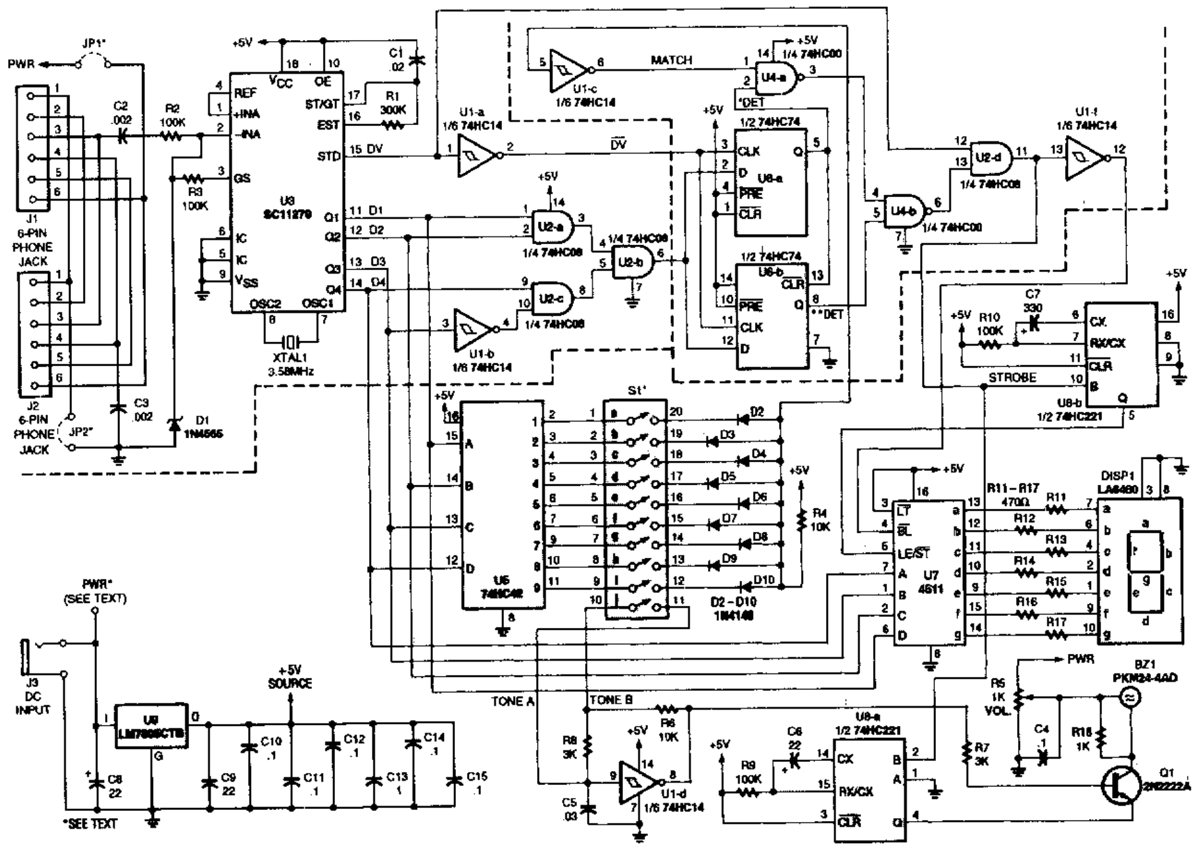


POPULAR ELECTRONICS

FIG. 100-13

Two hybrids (T1 and T2) are used to allow direct connection to a telephone line. This circuit uses the common speech-inversion algorithm where the frequency of an audio signal is inverted about a center frequency. An LM1496 balanced modulator is used to heterodyne the speech range against a 3.58-kHz signal.

PHONE PAGER

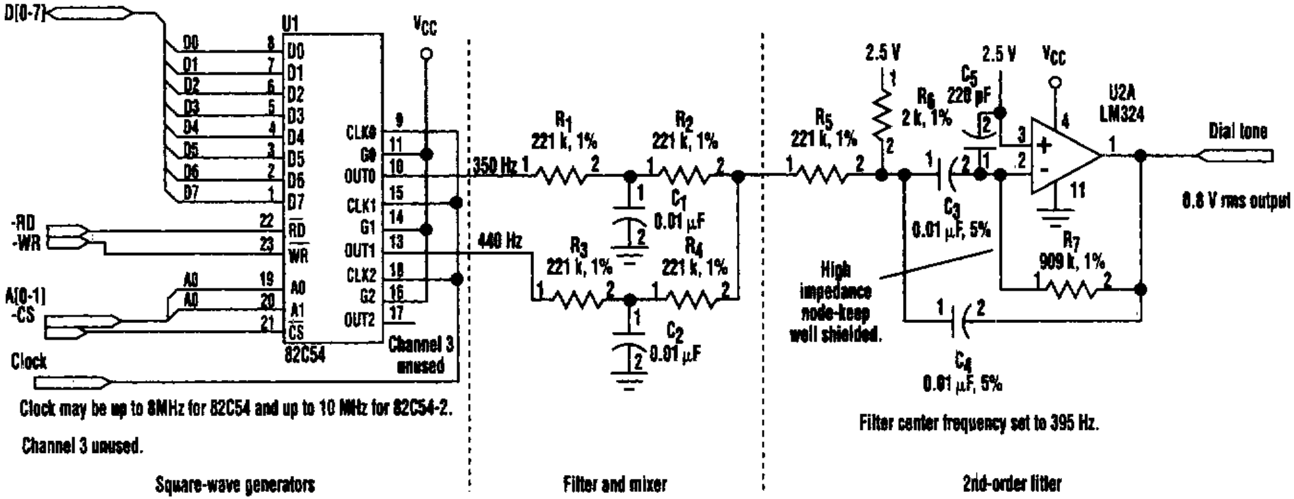


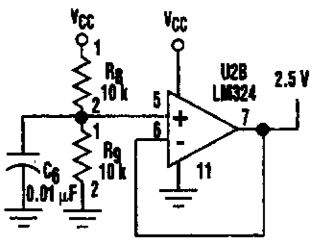
POPULAR ELECTRONICS

FIG. 100-14

This pager allows you to use your in-house phone wiring as a PA system. It uses two tone decoders to detect a particular touch-tone key. This key enables an audio amplifier.

5-V DIAL-TONE CIRCUIT






```
                  82C54 PROGRAMMING INFORMATION
---------------------------------------------------------------------
OUT BASE, 76h                   ; Set up channel 1 as sqr wave divider
OUT BASE+1, DIVISOR low byte    ; Enter divisor for 350Hz, low byte
OUT BASE+1, DIVISOR high byte   ; Enter divisor for 350Hz, high byte
OUT BASE, 0b6h                  ; Set up channel 2 as sqr wave divider
OUT BASE+2, DIVISOR low byte    ; Enter divisor for 440Hz, low byte
OUT BASE+2, DIVISOR high byte   ; Enter divisor for 440Hz, high byte

    For 1.8432MHz Clock, 350 Hz divisor = 5266 or 1492 hex.
    For 1.8432MHz Clock, 440 Hz divisor = 4189 or 105d hex.
```

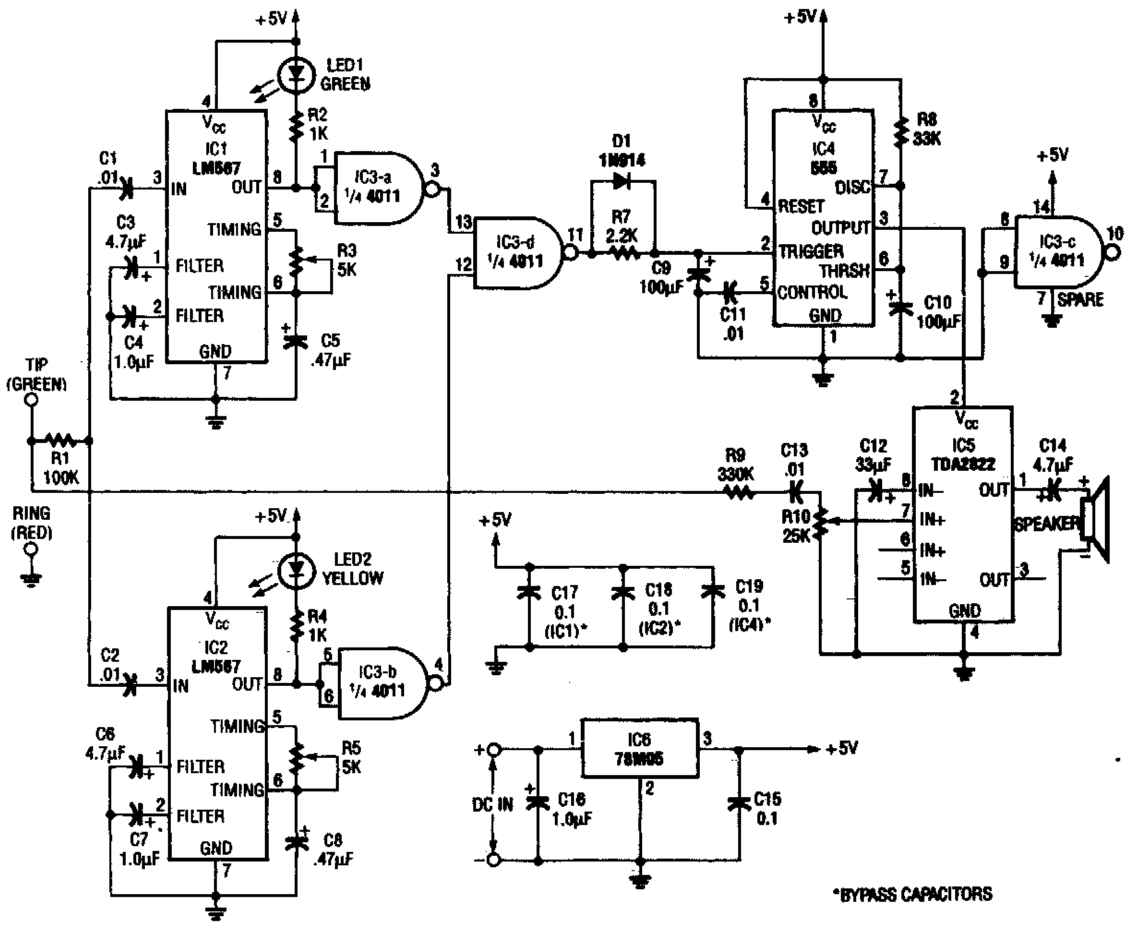
ELECTRONIC DESIGN

FIG. 100-15

This circuit uses inexpensive, common components to generate a precise dial tone for phone applications (see the figure). U1 (an Intel 82C54 timer-counter) generates 350- and 440-Hz square waves that are filtered by R_1/C_1 and R_3/C_2, and mixed together by resistors R2 and R4.

An operational amplifier configured as a 395-Hz, Sallen-Key, second-order bandpass filter (halfway between 350 and 440 Hz) removes unwanted signal harmonics. Almost any timer-counter can be used as the signal source, so long as it produces roughly square-wave outputs.

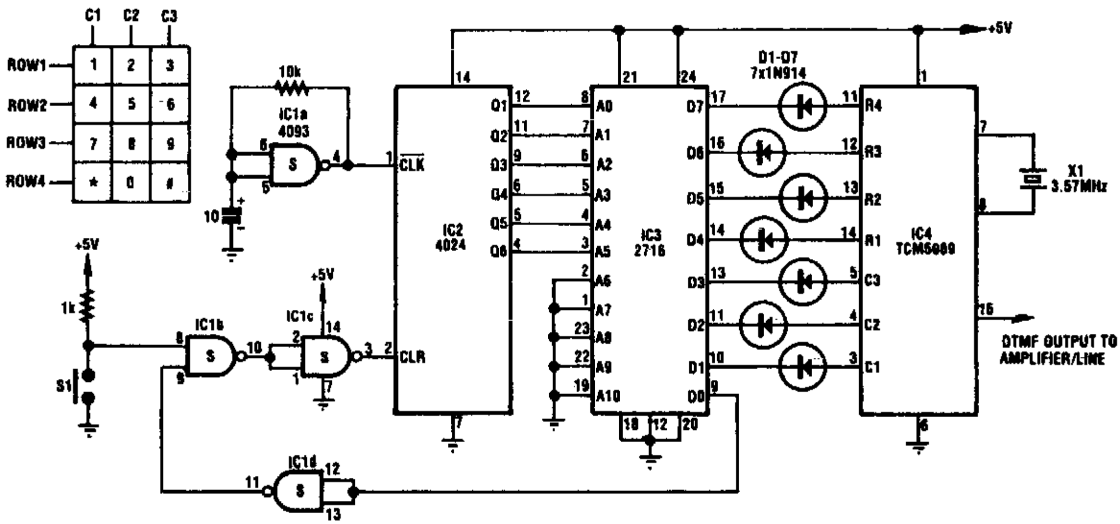
PHONE PAGER



ELECTRONICS NOW ***FIG. 100-16***

This pager works with DTMF phones. It displays a number and sounds an alert as the number on the display corresponds to a specific message.

ALARM DIALER

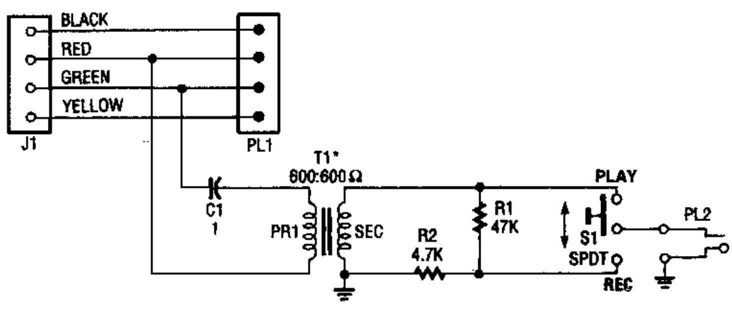


SILICON CHIP *FIG. 100-17*

This circuit dials a stored DTMF tone sequence from EPROM when a control line is taken to 0 V. IC1 is a Schmitt trigger oscillator, running at around 2 Hz. It clocks a 4024 binary counter. The counter's outputs connect to the address leads of the EPROM. A 2716 was used here, but the choice of EPROM is by no means critical.

Normally, the counter is held reset by a logic 1 on its reset pin (pin 2). When the trigger input is sent low, pin 10 of IC1 goes low, pin 3 goes high, and the reset is removed from the counter. It then begins to clock, incrementing the EPROM. When moved from address 000000, the data on bit D0 of the EPROM changes to a logic 1 and holds the circuit running. The last address should have data 11111110 to reset the circuit to standby.

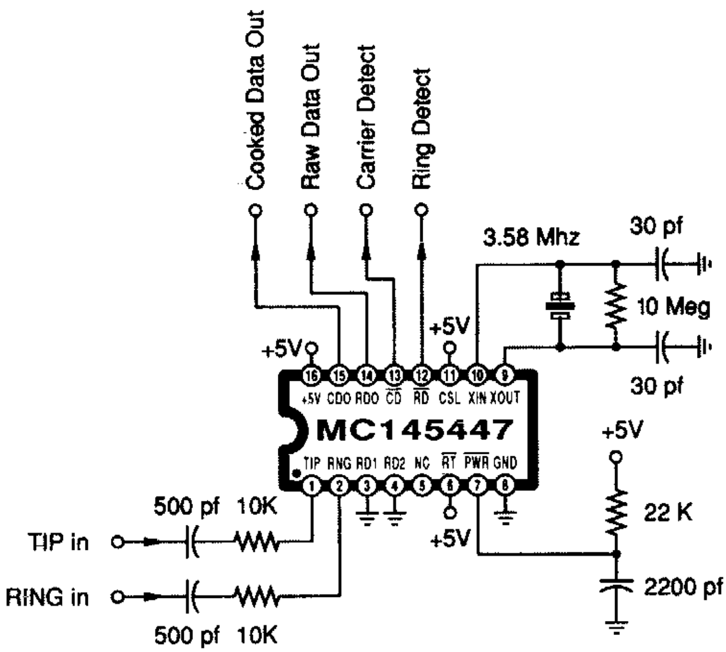
TELEPHONE AUDIO INTERFACE



POPULAR ELECTRONICS *FIG. 100-18*

Used to record and play back tapes via the phone lines, this simple circuit has an audio level switch (S1).

CALLER ID CIRCUIT

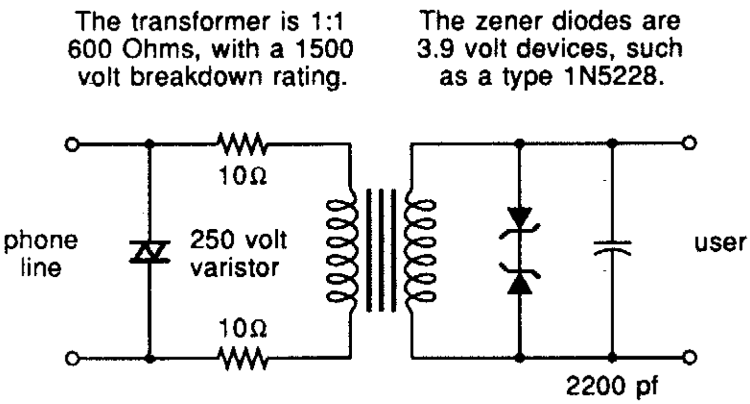


RADIO-ELECTRONICS

FIG. 100-19

This caller ID circuit uses the Motorola MC145447 IC chip. This service must be available from your local phone company in order for this circuit to be used.

FCC PART 68 PHONE INTERFACE

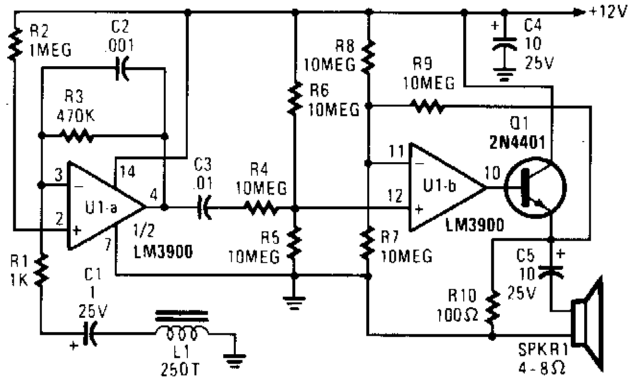


RADIO-ELECTRONICS

FIG. 100-20

An FCC Part 68 interface is required any time you connect any circut of your own to the phone line.

TELEPHONE AMPLIFIER

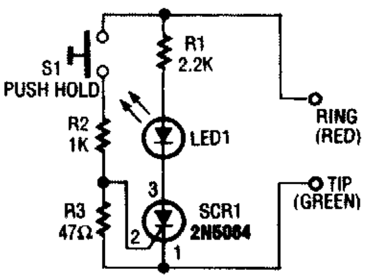


POPULAR ELECTRONICS **FIG. 100-21**

Section U1-a is configured as a high-gain inverting voltage amplifier that is inductively coupled to the phone line via L1. Inductor L1 is a homemade unit that consists of 250 turns of fine, enamel-coated wire that is wound on an iron core. The op amp receives the few mV produced by L1 via C1 and R1 and amplifies the signal. Capacitor C1 acts as the negative-feedback component that limits the circuit's high-frequency gain, while R3 limits the low-frequency gain. Resistor R3 is particularly important because without it, the amplifier would saturate.

Op amp U1-b is configured as a difference amplifier. It receives a signal from U1-a via C3 and R4 and amplifies the difference between it and half of the supply voltage. Transistor Q1 is configured as a common-collector amplifier ensuring sufficient signal to drive the speaker. Capacitor C5 is used to remove any dc component provided by transistor Q1.

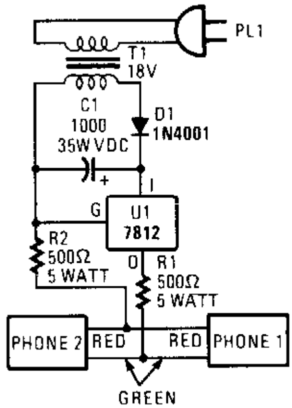
TELEPHONE HOLD CIRCUIT



When S1 is pressed, the SCR fires, and places LED1 and R1 across the phone line. The line voltage drops to about 20 V, which holds the connection to the phone company's central office.

ELECTRONICS NOW **FIG. 100-22**

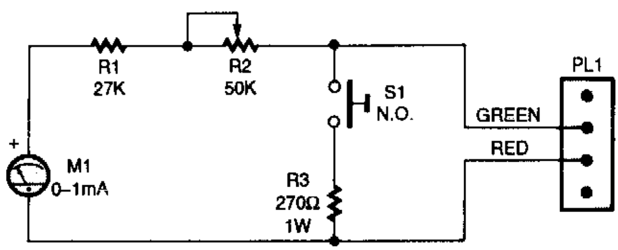
TELEPHONE CIRCUIT



POPULAR ELECTRONICS ***FIG. 100-23***

This circuit is useful for checking out old telephones by providing them with the dc voltage that they require for operation.

TELEPHONE-LINE TESTER



POPULAR ELECTRONICS ***FIG. 100-24***

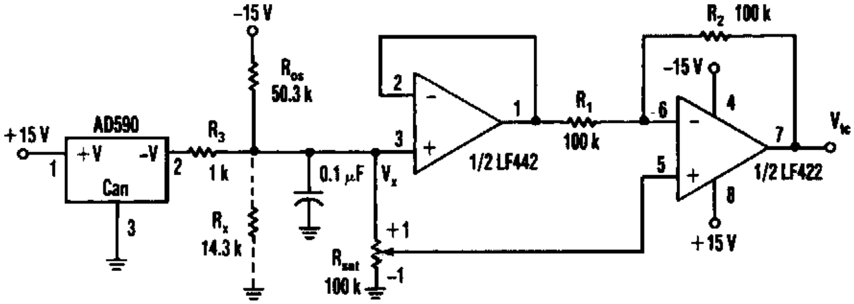
The telephone-line tester consists of nothing more than a meter (that's used to measure line voltage in the on- and off-hook state), three resistors (one of which is variable), a pushbutton switch, and a modular telephone connector. When the circuit is connected to the telephone line, a meter reading of 5 to 10 V (when S1 is pressed) indicates that the line is okay.

101

Temperature-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

TEMPERATURE COMPENSATION ADJUSTER

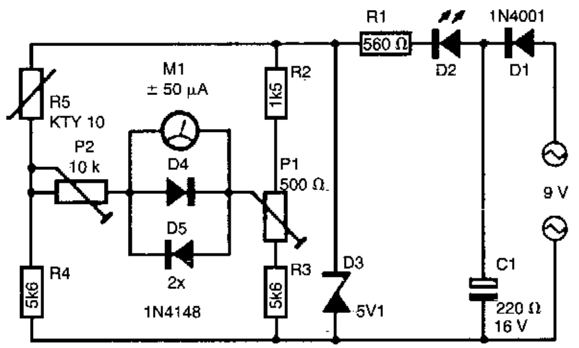


ELECTRONIC DESIGN

FIG. 101-1

The circuit shown delivers +10 to −10 mV°/C output using an Analog Devices' AD590 temperature transducer. R_x is a scaling resistor.

THERMOMETER FOR 5-V OPERATION



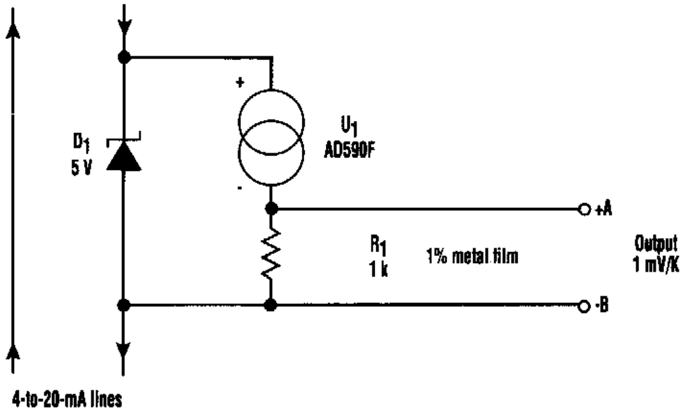
303 CIRCUITS

FIG. 101-2

At the heart of this simple circuit is the well-known type KTY10 temperature sensor from Siemens. This silicon sensor is essentially a temperature-dependent resistor that is connected as one arm in a bridge circuit here. Preset P1 functions to balance the bridge at 0°C. At that temperature, moving coil meter M1 should not deflect, i.e., the needle is in the center position. Temperature variations cause the bridge to be unbalanced, and hence produce a proportional indication on the meter. Calibration at, say, 20°C is carried out with the aid of P2.

The bridge is fed from a stabilized 5.1-V supply, based on a temperature-compensated zener-diode. It is also possible to feed the thermometer from a 9-V battery, provided D1–D3, R1 and C1 are replaced with a Type 78L05 voltage regulator, because this is more economic as regards to current consumption.

HOOK SENSOR ON 4- TO 20-mA LOOP

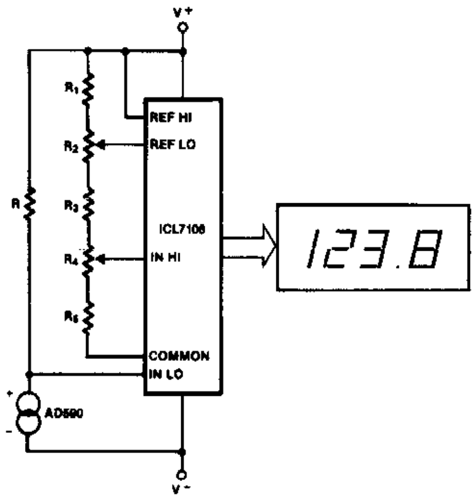


ELECTRONIC DESIGN **FIG. 101-3**

Here's an effective for a temperature sensor to receive power from a 4-to-20 mA loop without actually affecting the loop current (see the figure). This particular temperature sensor IC (AD590F) conducts 1 μA/K when powered by a supply in the range of 4 V to 40 Vdc.

The scheme uses a 5-V Zener diode (D1) to regulate the power source for AD590F. Most of the current flows through the Zener diode and a small current flows through AD590F. A high-impedance device can read the temperature information across R1, which is a 1 mV/K in the range of –55°C to 150°C. The waste of power is negligible in this arrangement.

BASIC DIGITAL THERMOMETER



	R	R_1	R_2	R_3	R_4	R_5
°F	9.00	4.02	2.0	12.4	10.0	0
°C	5.00	4.02	2.0	5.11	5.0	11.8

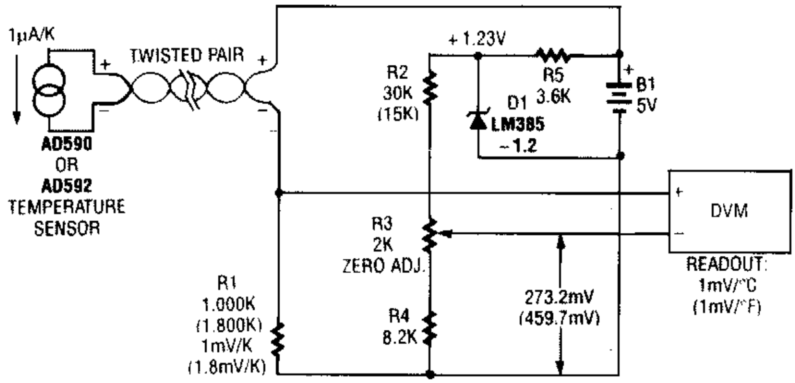
$\sum_{n=1}^{5} R_n = 28k\Omega$ nominal

All values in kΩ

The ICL7106 has a V_{IN} span of ±2.0V, and a V_{CM} range of (V^+ −0.5) Volts to (V^- +1) Volts; R is scaled to bring each range within V_{CM} while not exceeding V_{IN}. V_{REF} for both scales is 500mV. Maximum reading on the Celsius range is 199.9°C, limited by the (short-term) maximum allowable sensor temperature. Maximum reading on the Fahrenheit range is 199.9°F (93.3°C), limited by the number of display digits. See note next page.

INTERSIL **FIG. 101-4**

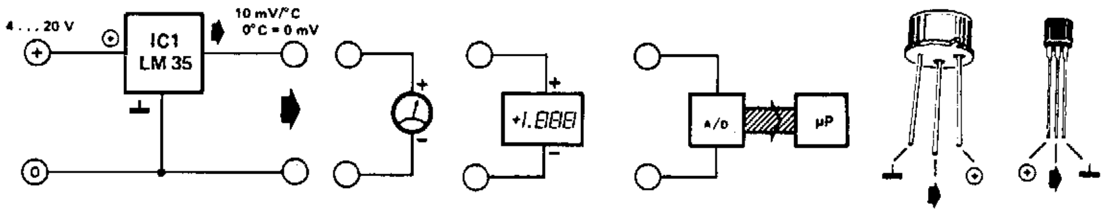
REMOTE TEMPERATURE SENSING



RADIO-ELECTRONICS **FIG. 101-5**

An AD590 or AD592 makes it easy to transmit temperature data over a pair of wires. The circuit produces 1mV/°C (or 1mV/°F using the values in parentheses).

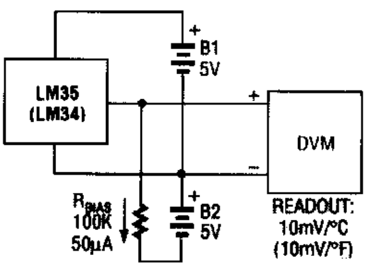
TEMPERATURE SENSOR



303 CIRCUITS **FIG. 101-6**

The LM35 temperature sensor provides an output of 10 mV/°C for every degree Celsius over 0°C. At 20°C the output voltage is 20 × 10 = 200 mV. The circuit consumes 60 μA. The load resistance should not be less than 5 kΩ. A 4- to 20-V supply can be used.

LOW TEMPERATURE SENSOR



A negative bias current can produce the offset needed for below-zero readings using the LM34 or LM35 temperature sensor.

RADIO-ELECTRONICS **FIG. 101-7**

ELECTRONIC THERMOSTAT

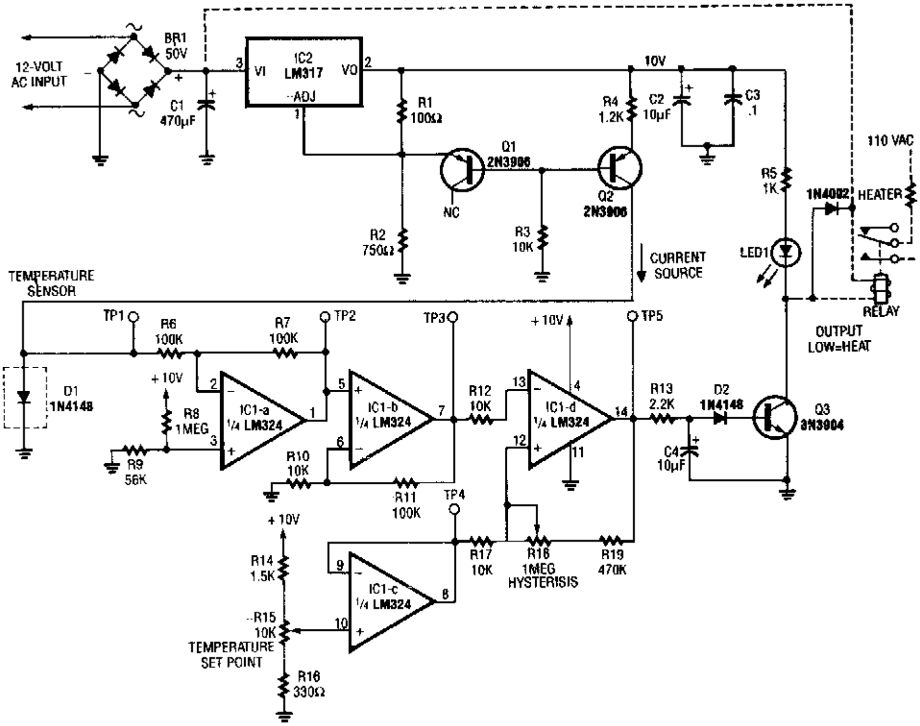


TABLE 1—RESISTOR VALUES

Temperature Range (Degrees C)	R14	R15	R16
−50 to −30	10K	1K	330Ω
−30 to −10	9.1K	1K	1.2K
−10 to 15	8.2K	1K	2.2K
15 to 35	7.5K	1K	3.3K
35 to 55	6.2K	1K	4.3K
55 to 75	5.1K	1K	5.1K
75 to 95	4.3K	1K	6.2K
95 to 115	3.3K	1K	6.8K
115 to 135	2.2K	1K	8.2K
135 to 155	1.2K	1K	9.1K

RADIO-ELECTRONICS *FIG. 101-8*

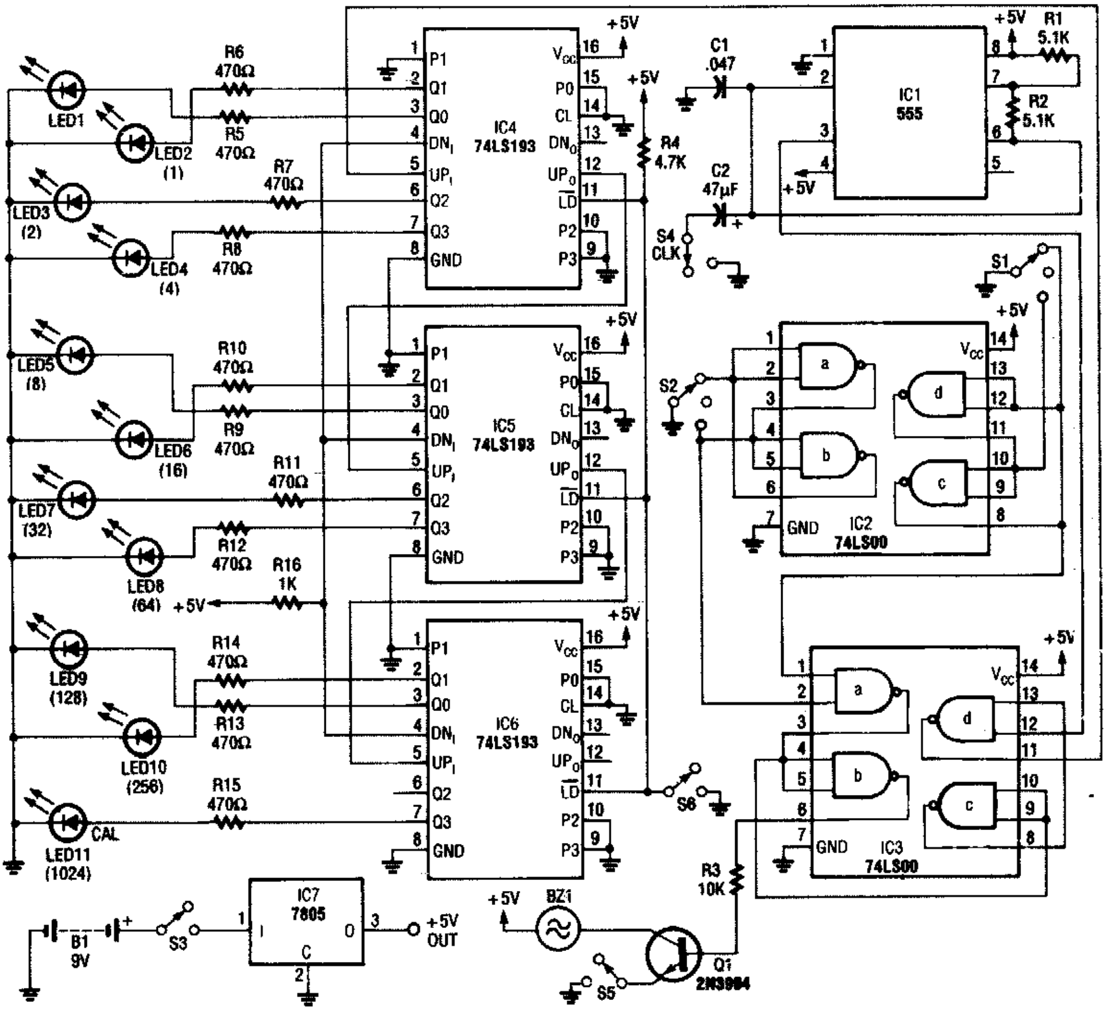
A diode, such as a IN4148, has a typical –2m V/°C temperature coefficient at a 1 mA diode current. Q1 and Q2 form a constant current source. D1 is the temperature sensor. IC1-a and -b are dc amplifiers, with IC1-c a temperature reference voltage supply. IC1-d is a comparator with variable hysteresis. R14, R15, and R16 are chosen depending on the thermostat range desired. Q3 is a relay driver (2N3904). The relay used should handle the load current or an optoisolator triac combination can be used.

102

Timer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

REFLEX TIMER

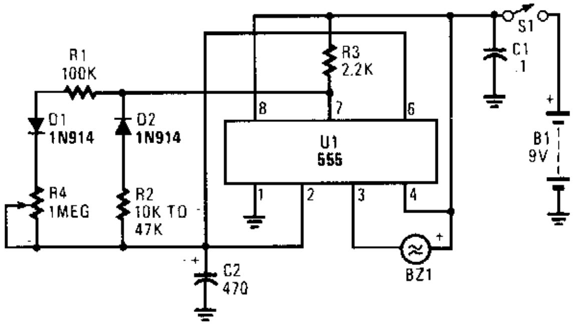


ELECTRONICS NOW

FIG. 102-1

This timer circuit uses a 555 IC timer and three 74LS193 counters to drive an LED display. S1 is activated by one person, who turns on piezo buzzer BZ1 via Q1 and also starts the clock; S1 is activated by the other person being timed. This shuts off the timer, and the number of LEDs lit indicate, in binary form, the elapsed time.

TELE-TIMER



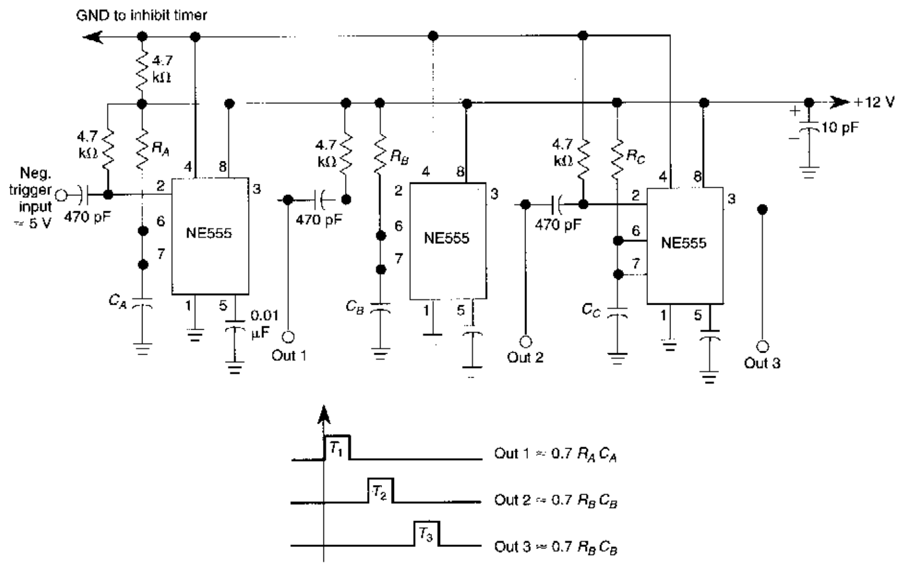
POPULAR ELECTRONICS ***FIG. 102-2***

The circuit is built around a 555 oscillator/timer. The circuit provides two time periods. The long-running time period is adjustable from about 1 to 10 minutes, and the short time period is preset to about three seconds.

Here's how the dual timer operates. When the power is switched on, C2 begins to charge through R3, R1, D1, and R4 to start the long-term timer period. When the voltage across C2 reaches the 555's internal switching point, the long-term timer times out, discharging C2 through R2, D2, and pin 7 of the 555. During that time, pin 3 of the 555 is pulled to ground, activating the piezo sounder.

To set the short time period to about four seconds, use a 10 k resistor for R2, and for about twenty seconds use a 47 k resistor. The timing capacitor, C2, should be a good-quality, low-leakage unit.

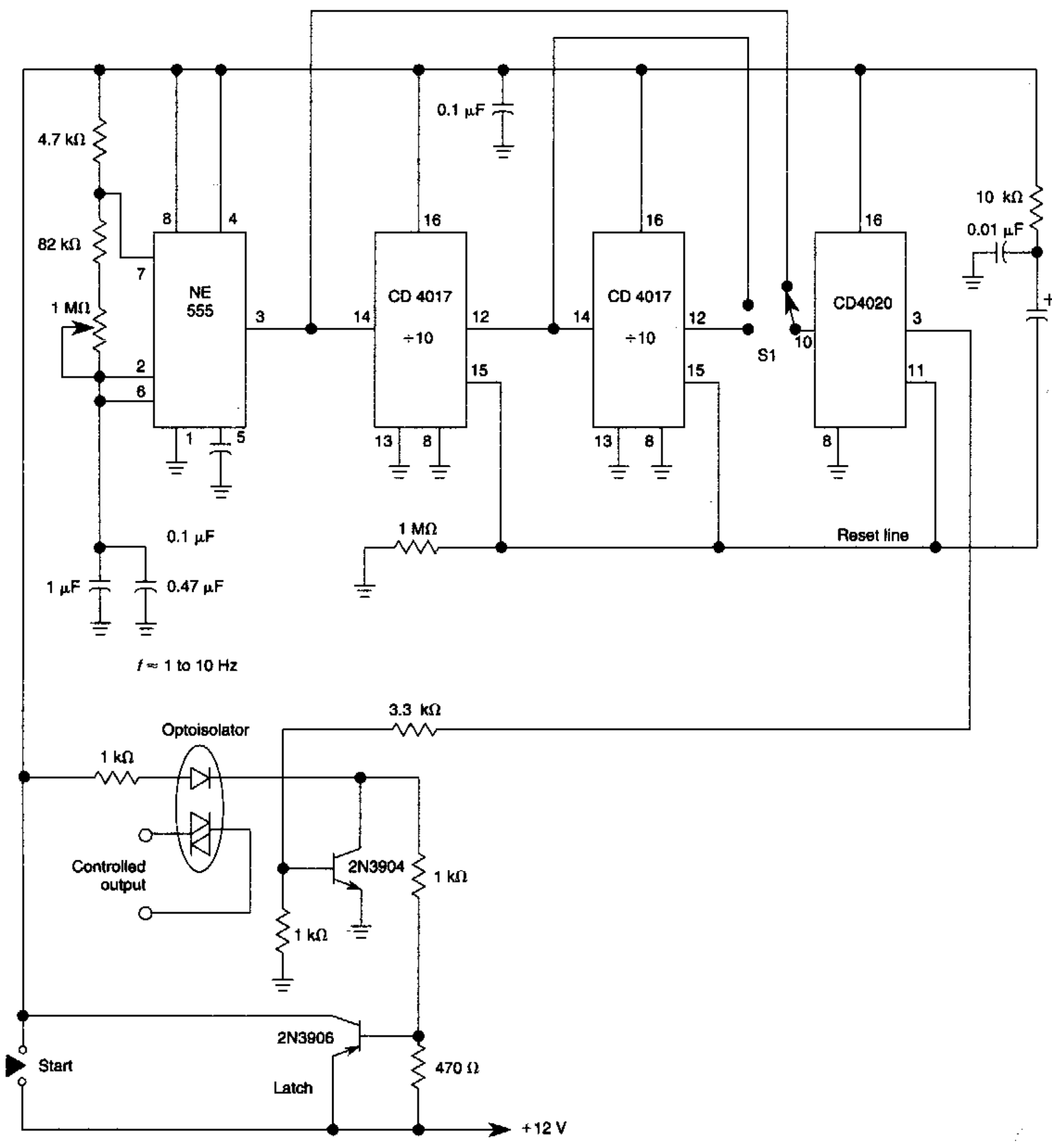
THREE-STAGE SEQUENTIAL TIMER



WILLIAM SHEETS ***FIG. 102-3***

By using three 555 ICs, three sequential pulses can be generated. Output 3 can be connected back to trigger input to achieve astable operation.

WIDE-RANGE TIMER–1MINUTE TO 400 HRS

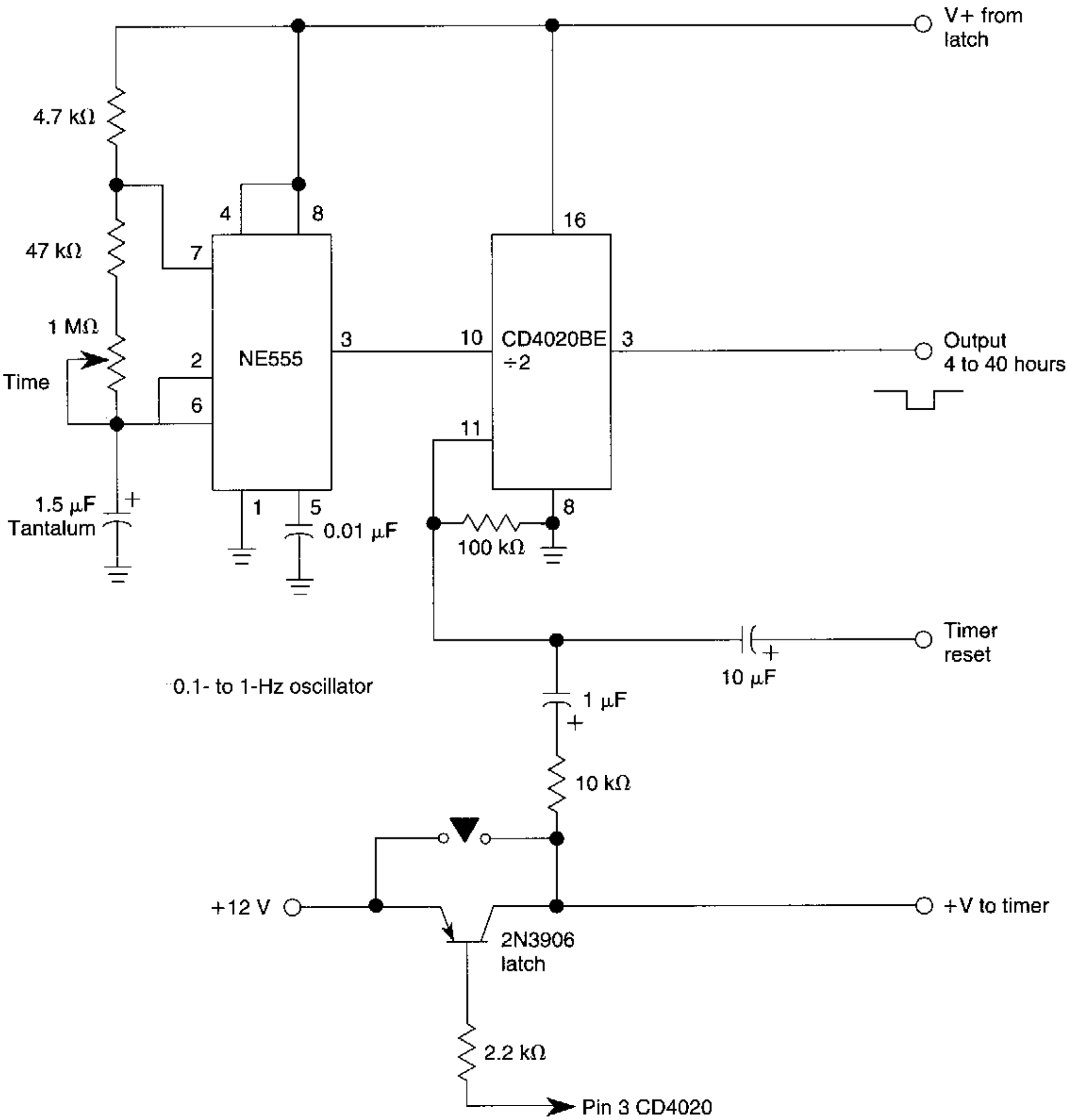


WILLIAM SHEETS

FIG. 102-6

This ultra wide range timer uses a 555 timer base, two 4017Bs and a 4020B that act as frequency dividers that can be switched in and out. S1 is a SP3T range switch.

LONG-DELAY-PERIOD TIMER

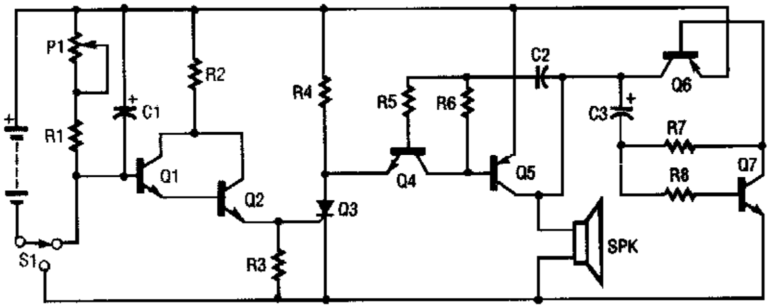


WILLIAM SHEETS

FIG. 102-7

This method of obtaining a 4 to 40 hour timing period from a 555 IC can be further expanded to produce even longer delays with equal accuracy.

COUNT-DOWN TIMER

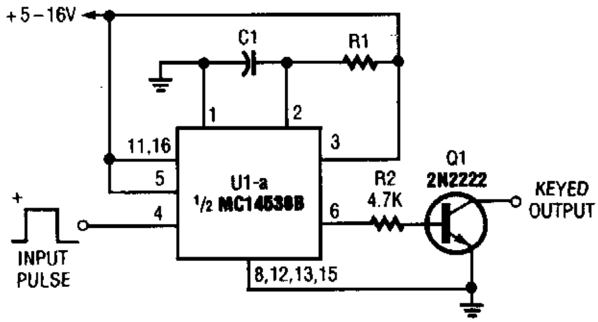


1991 PE HOBBYIST HANDBOOK *FIG. 102-8*

C1	100-µF Electrolytic Capacitor	R3	33-kΩ Resistor
C2	0.0047-µF Mylar Capacitor	R4	200 Ω Resistor
C3	1-µF Electrolytic Capacitor	R5	2.2-kΩ Resistor
P1	2-MΩ Trimmer Resistor	R6	220-kΩ Resistor
Q1, Q2, Q4, Q7	2N3904 Transistor	R7	2.2-MΩ Resistor
Q3	106 SCR	R8	7.5-kΩ Resistor
Q5, Q6	2N3906 Transistor	S1	SPDT Slide Switch
R1	1-MΩ Resistor	SPK	Small Speaker
R2	10-kΩ Resistor	Misc	PC Board, 9-V Snap Wire

With switch S1 in the off position, as shown, battery voltage is applied across timing-capacitor C1, which stays charged while the rest of the circuitry has no power supplied to it. Transistor Q1, and thus transistors Q2 through Q4, are kept in an off condition as long as C1 has a sufficient charge.

EXTENDED ON-TIME TIMER



POPULAR ELECTRONICS *FIG. 102-9*

Half of a Motorola MC14538B dual, precision, retriggerable monostable multivibrator is used to form an extended on-time timer circuit. That type of circuit can be used as a switch debouncer. Such circuits are often used in digital circuitry, where each and every bounce of a switch contact is seen as a separate digital input.

The delay on time (established by C1 and R1) is easily set using the formula, $C_1 \times R_1 = T$, where C_1 is in microfarads, R_1 is in megohms, and T is in seconds.

103

Tone Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Repeater-Tone Burst Generator
Two-Tone Encoder

REPEATER-TONE BURST GENERATOR

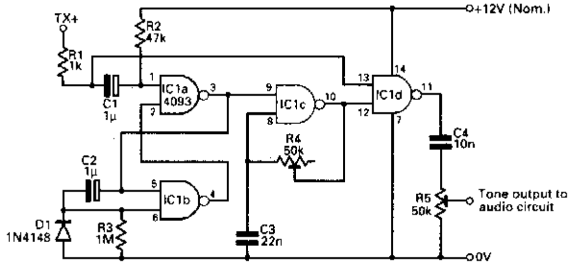


Fig. 1: The circuit, based on a single c.m.o.s chip and a few other components.

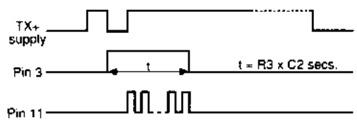


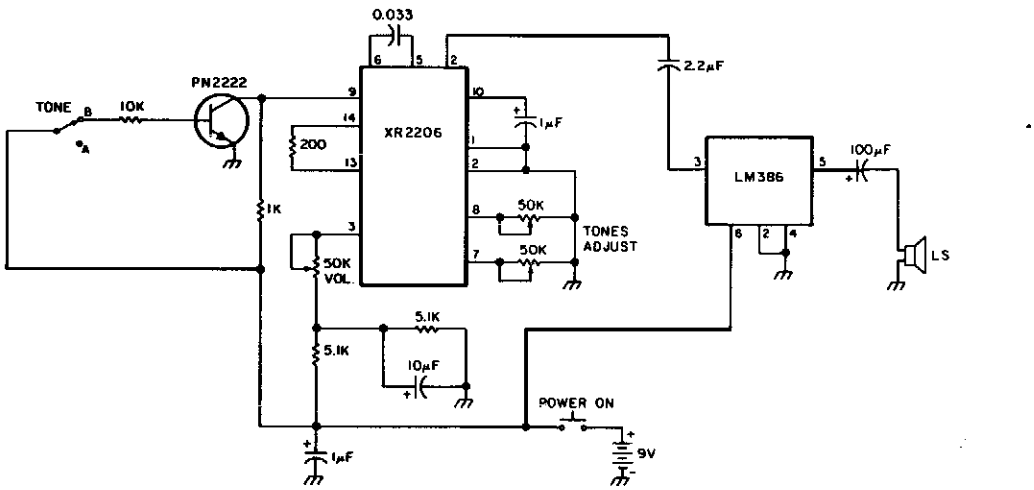
Fig. 2: Pulse and timing diagram, see the text for more details.

PRACTICAL WIRELESS

FIG. 103-1

Integrated circuit gates IC1-a and IC1-b form a monostable, whose time constant is determined by C2 and R3. When the transmitter is dekeyed (and then almost immediately rekeyed) point TX+ goes low and takes pin 1 low for a short time. This triggers the start of the timing period controlled by C_2/R_3. The capacitor C2, charges via R3 until the trigger point of gate IC1-b is reached. At this point, the monostable changes state and pin 3 goes low again. On the prototype, this time was about 700 ms. The pulse occurs each time after dekeying and it is normally inaudible. If, however, point TX+ goes high again (as in immediate rekeying) the monostable is still in the enabled state and the oscillations of IC1-c are present in the transmission. During this time period, the buffer gate, IC1-d, is enabled and the tone is therefore passed to the output.

TWO-TONE ENCODER



73 AMATEUR RADIO

FIG. 103-2

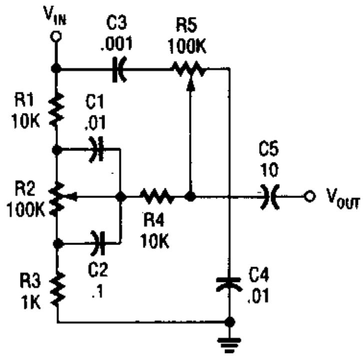
Using an XR2206 oscillator, this circuit can generate two audio tones. Switching between tones can be done with a logic level to either the base of the PN2222 or pin 9 of the XR2206.

104

Tone-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

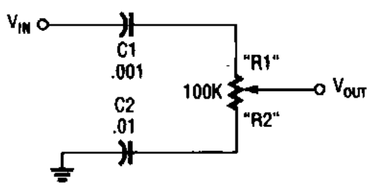
COMBINED BASS AND TREBLE CONTROLS



1993 ELECTRONICS HOBBYIST HANDBOOK ***FIG. 104-1***

Bass and treble circuits can be combined to form a two-control tone-adjust circuit, as shown here.

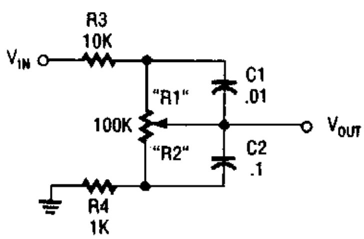
TREBLE TONE CONTROL



1993 ELECTRONICS HOBBYIST HANDBOOK ***FIG. 104-2***

The treble control has capacitors placed in series with the potentiometer.

BASS TONE CONTROL



1993 ELECTRONICS HOBBYIST HANDBOOK ***FIG. 104-3***

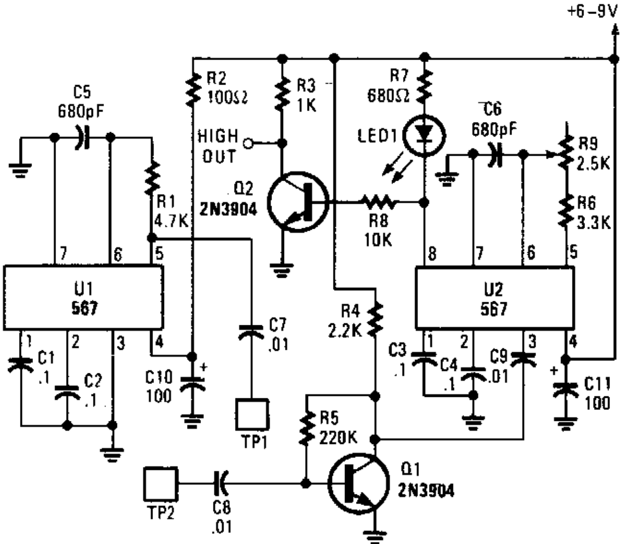
The frequency dependence of the capacitor's impedance permits this circuit to boost the bass frequencies.

105

Touch-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

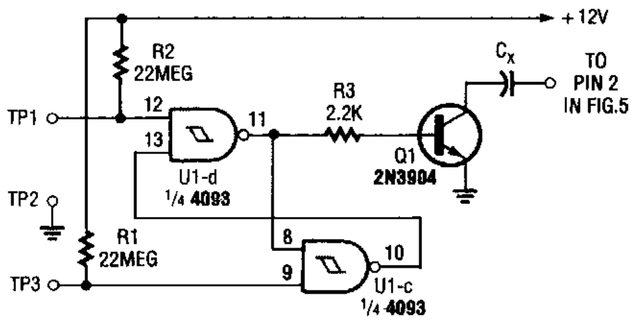
BRIDGING TOUCH PLATE SENSOR



POPULAR ELECTRONICS ***FIG. 105-1***

In this circuit, two 567 tone decoders are used. One is an oscillator, the other is a detector. Bridging TP1 and TP2 causes U2 to receive U1's signal, which causes pin 8 of U2 to go low. This action lights LED1 and drives the output of Q2 high.

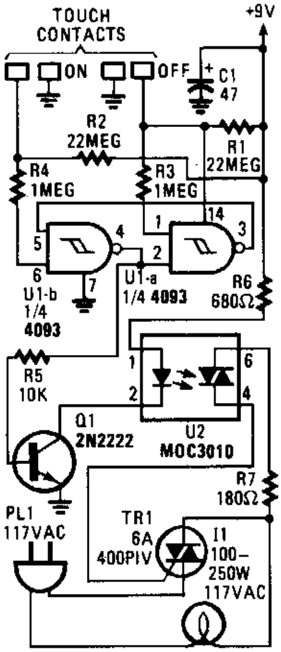
TOUCH SWITCH I



POPULAR ELECTRONICS ***FIG. 105-2***

Two NAND Schmitt triggers are connected in a flip-flop configuration to produce a bridged touch-activated switch.

TOUCH SWITCH II

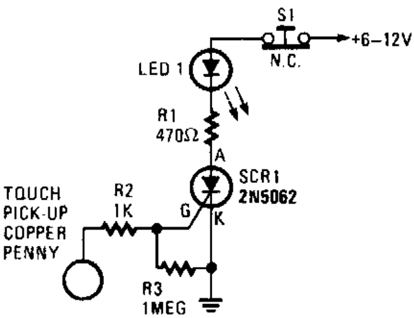


When the touch-on contacts are bridged, pin 6 of U1-b goes low, which forces its output (the set output) at pin 4 to go high. That high divides along two paths: in one path, the output is applied to pin 2 of U1-a, which causes its output at pin 3 to go low. That low is, in turn, applied to pin 5 of U1-b, which latches the gate in a high output state. In the other path, the output of U1-b is used to drive Q1. When Q1 turns on, U2's internal LED lights, which turns on its internal, light-sensitive, triac-driver (diac) output element. The triac driver feeds gate current to TR1, causing it to turn on, and light the lamp (I1).

When the off contact is bridged, U1-a's output switches and latches high, causing U1-b's output to go low, turning off the lamp.

POPULAR ELECTRONICS *FIG. 105-3*

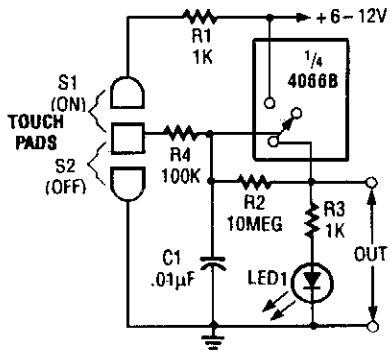
TOUCH ON-ONLY SWITCH



HANDS-ON ELECTRONICS *FIG. 105-4*

This touch on-only switch can be triggered into conduction by electrical means, and can only be reset by way of a mechanical switch. When the touch terminal is contacted by a finger, the SCR turns on and illuminates LED1.

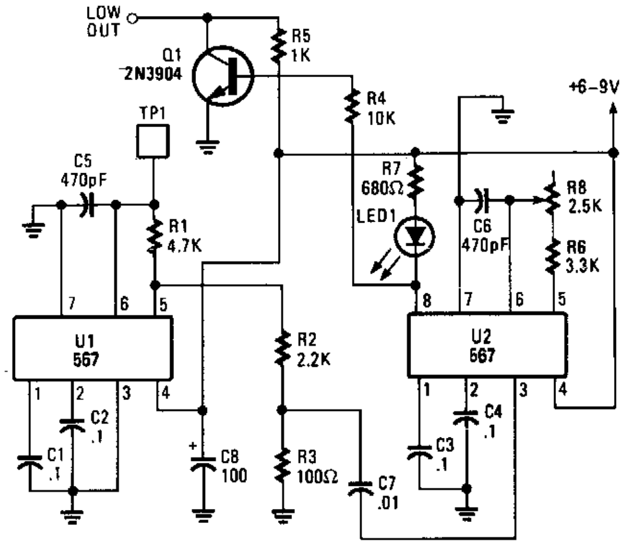
LATCHING TOUCH SWITCH USING CD4066B



RADIO-ELECTRONICS *FIG. 105-5*

When touch switch S1 is activated, R4 is driven high, and the control voltage goes high, which latches the switch. When S2 is activated, R4 goes low and the control voltage goes low, which deactivates the switch.

SINGLE-PLATE TOUCH SENSOR



POPULAR ELECTRONICS ***FIG. 105-6***

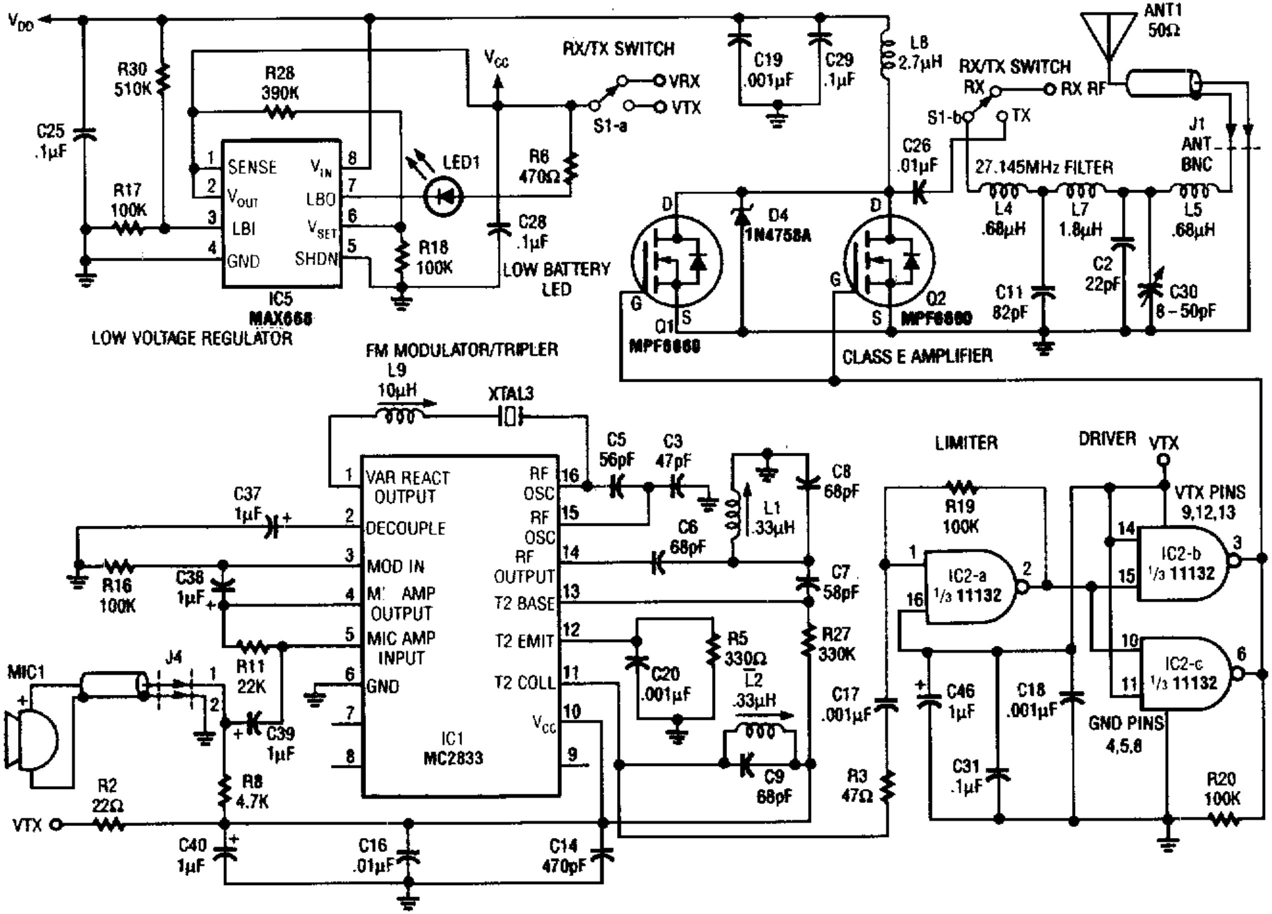
This system operates on the principle that capacitance loading of an oscillator will lower its frequency. When a foreign body comes into contact with touch plate, the frequency of U1 is lowered. This removes the oscillator signal from U1 from U2's passband, which causes U2 to lose lock, turns off the LED, and causes the collector of Q1 to go low.

106

Transmitter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

27.125-MHz NBFM TRANSMITTER

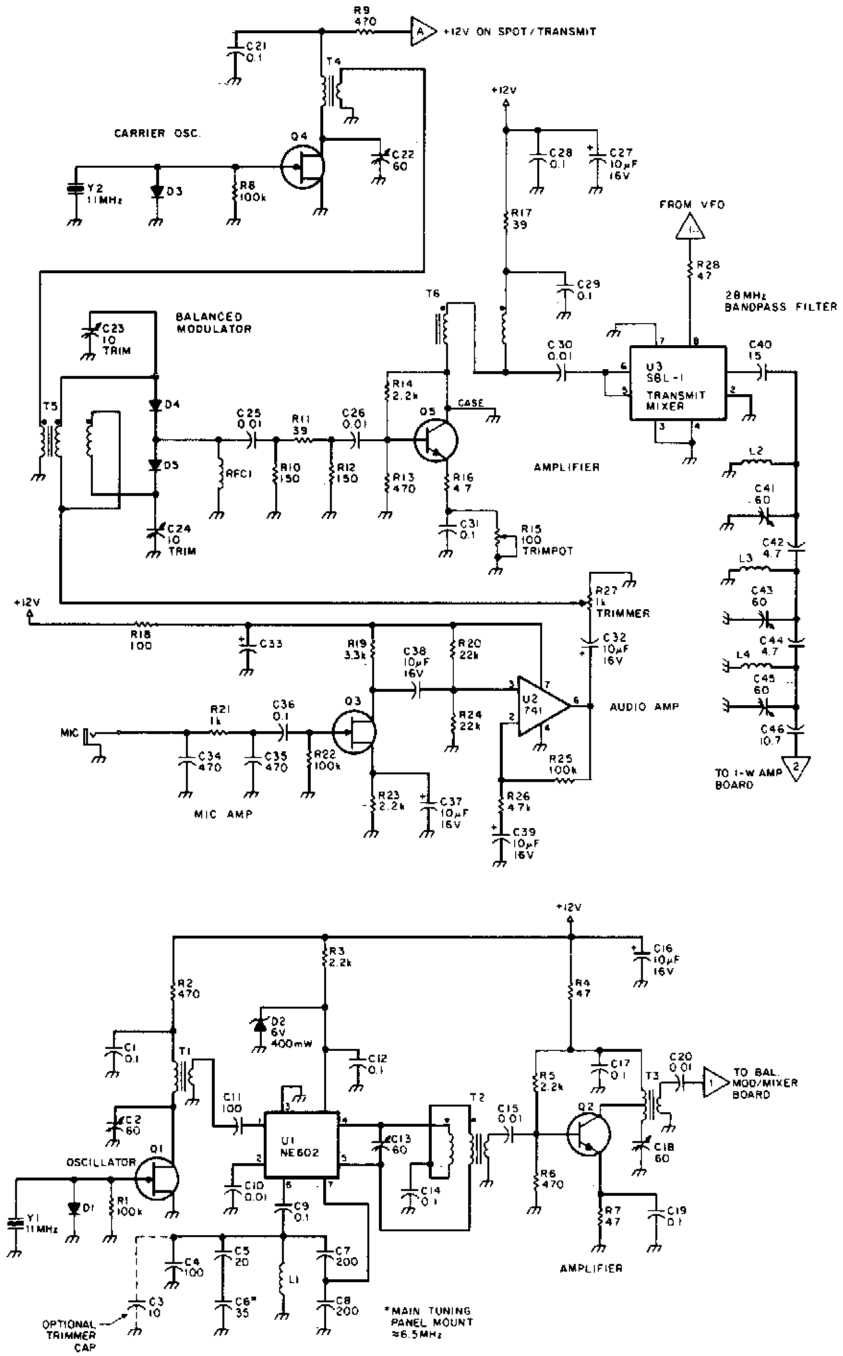


ELECTRONICS NOW

FIG. 106-1

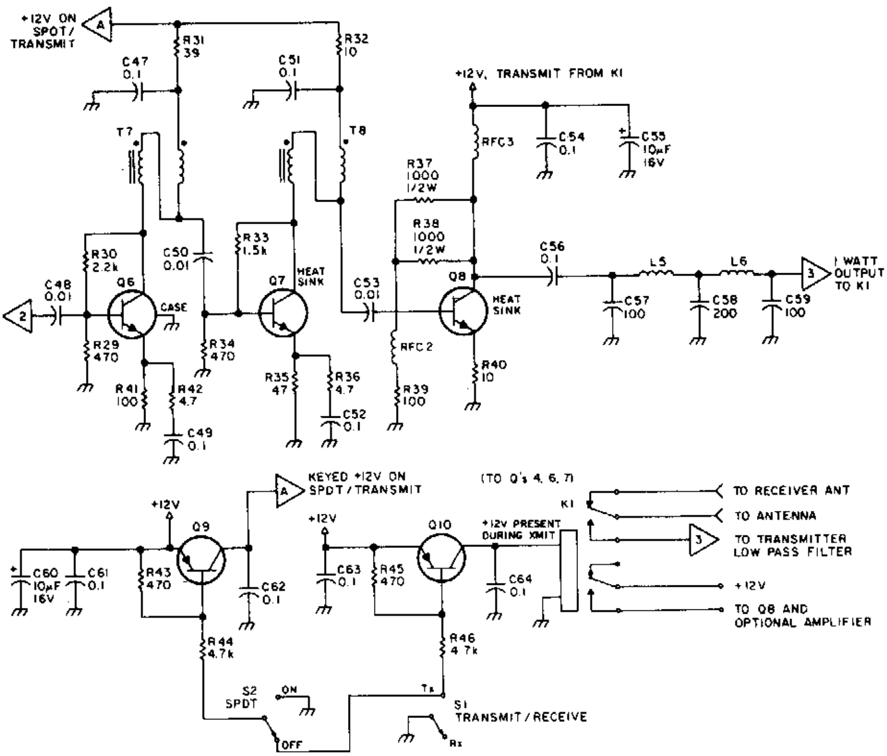
Using a Motorola MC2833 one-chip FM transmitter, a few support components, and an MPF6660 FET RF amp, this transmitter delivers about 3 W into a 50-Ω load. It is capable of operation over about 29 to 32 MHz with the components shown.

10-M DSB QRP TRANSMITTER WITH VFO



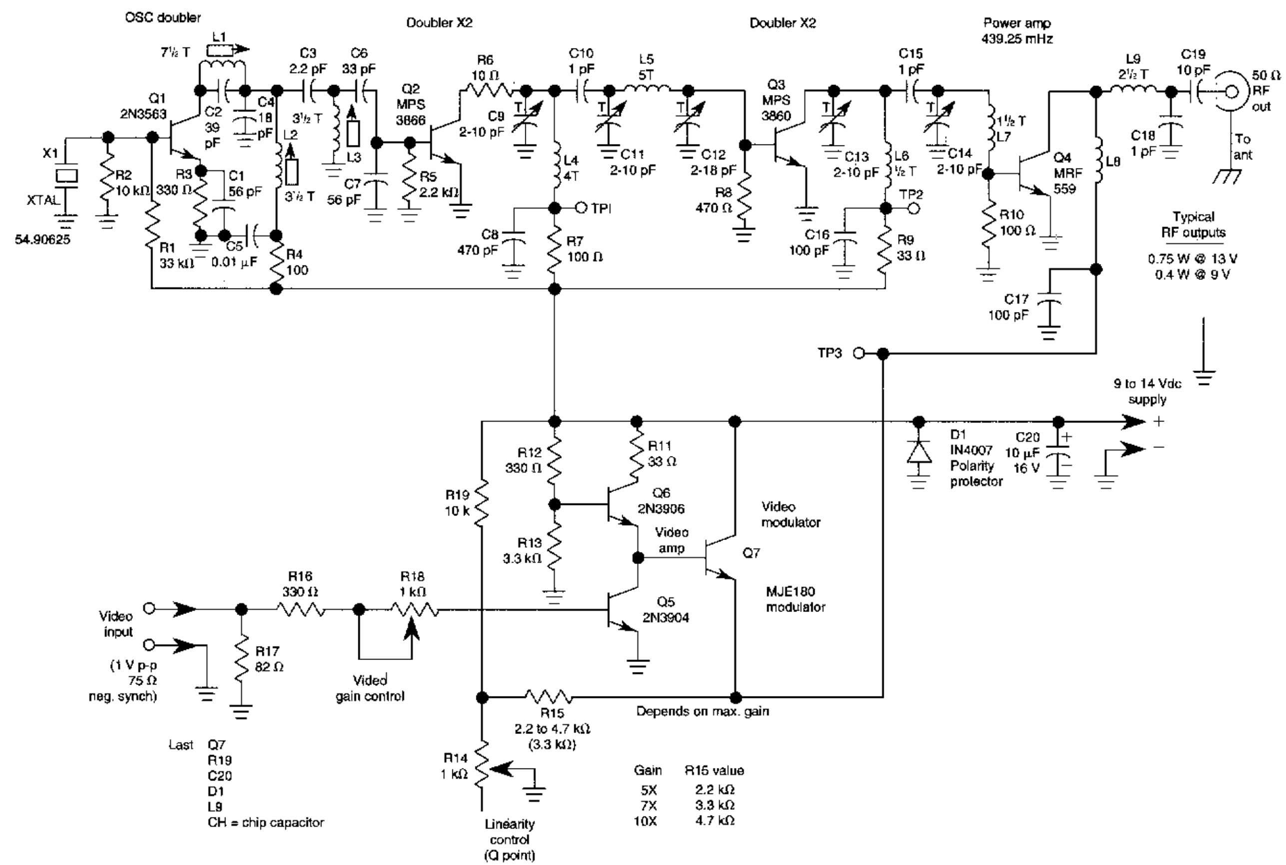
73 AMATEUR RADIO TODAY

FIG. 106-2



The three schematics represent three building blocks for a 10-meter SSB transmitter. Or these blocks can be used separately as circuit modules for other transmitters. The VFO board uses an FET transmittal oscillator, the VFO signal is mixed in an NE602 mixer and is amplified by Q2 to a level sufficient to drive an SBL-1 mixer in the transmit mixer stage (+7 to +10 dBm). In the balance mixer/modulator board, an 11-MHz crystal oscillator drives a diode balanced mixer. Audio for modulation purposes is also fed to this mixer. The DSB signal feeds a 28-MHz BPF. The 1-W amplifier board consists of a 3-stage amplifier and transmit/receive switching circuitry.

ATV JR TRANSMITTER 440 MHz

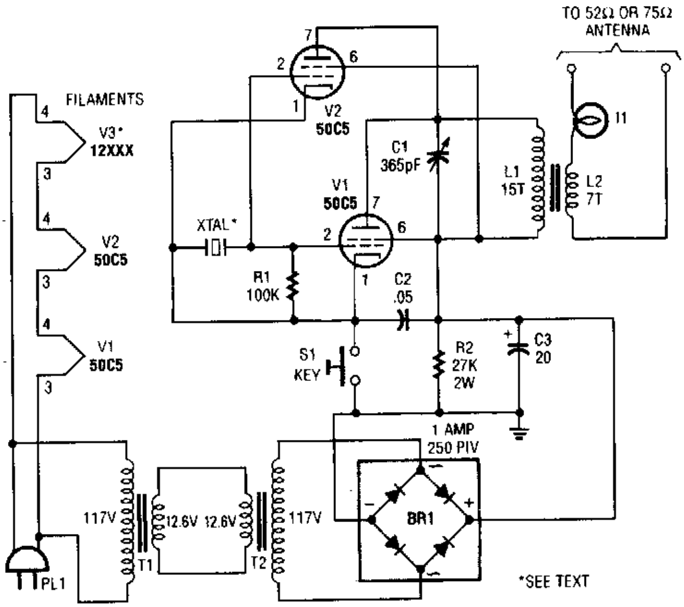


WILLIAM SHEETS

FIG. 106-3

This low-power video transmitter is useful for R/C applications, surveillance, or amateur radio applications. Seven transistors are used in a crystal oscillator-multiplier RF power amplifier chain, and a high-level video modulator. A 9- to 14-Vdc supply is required. Output is 0.4 to 1.2 W, depending on supply voltage. A complete kit of parts is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-00530

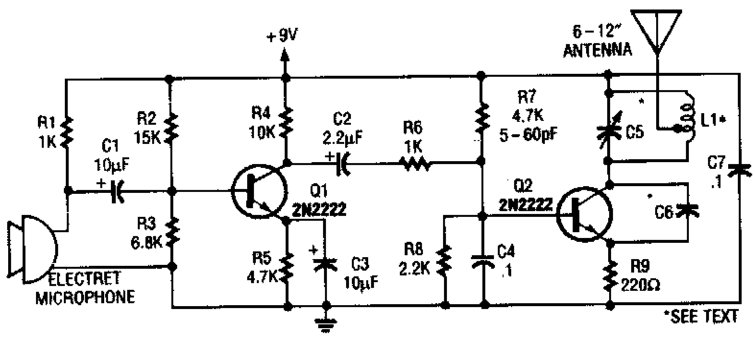
6-W ECONOMY MORSE-CODE TRANSMITTER FOR 7 MHz



POPULAR ELECTRONICS *FIG. 106-4*

The vacuum tube is still alive and useful in some applications, as in this CW transmitter. The circuit was built in old-fashioned breadboard style on a wooden base. Old table radios are a good source of parts for this circuit. V3 is used as a ballast resistor—a 75-Ω or 100-Ω 5-W resistor could be substituted. L1 is 15 turns of hookup wire on a ⅞" form 2" long. L2 is 7 turns of the same wire. L2 is wound over L1. Be careful as up to 160 V is present on V1 and V2.

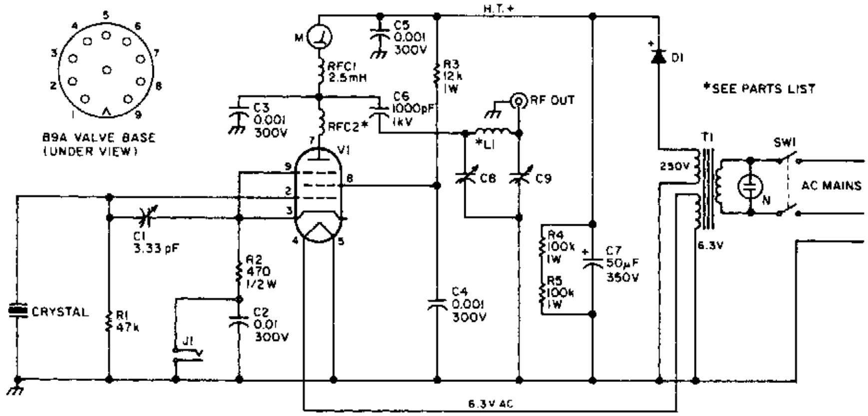
SIMPLE FM TRANSMITTER



RADIO-ELECTRONICS *FIG. 106-5*

Running from a 9-V battery, this transmitter can be used as a wireless microphone with an ordinary 88- to 108-MHz FM broadcast receiver. Keep the antenna length under 12 inches to comply with FCC limits. L1 is 6 turns of #24 wire wound around a pencil or a ¼" form, with turns spaced 1 wire diameter. C6 is a gimmick capacitor of about 1 pF.

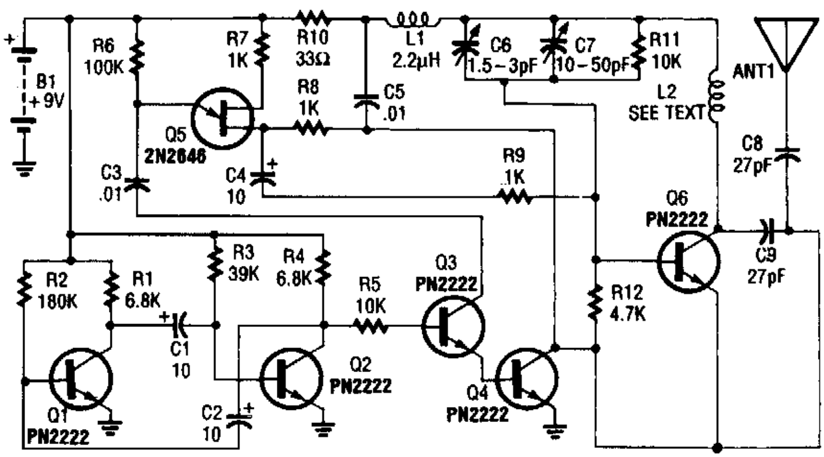
VACUUM-TUBE LOW-POWER 80/40-METER TRANSMITTER



73 AMATEUR RADIO TODAY ***FIG. 106-6***

Using a 6BW6 vacuum tube, the above transmitter delivers about 5 W output. C1 is adjusted for cleanest CW note. C8 and C9 are 365 pF and dual-365 pF (paralleled) tuning capacitors. L1 is 35 turns of #24 enamelled wire on a 1" plastic tube. FT-243 crystals for 3.5 or 7 MHz are used. Do not use this circuit to produce a 7-MHz output from a 3.5-MHz crystal—it is not intended to "double over" crystal frequencies.

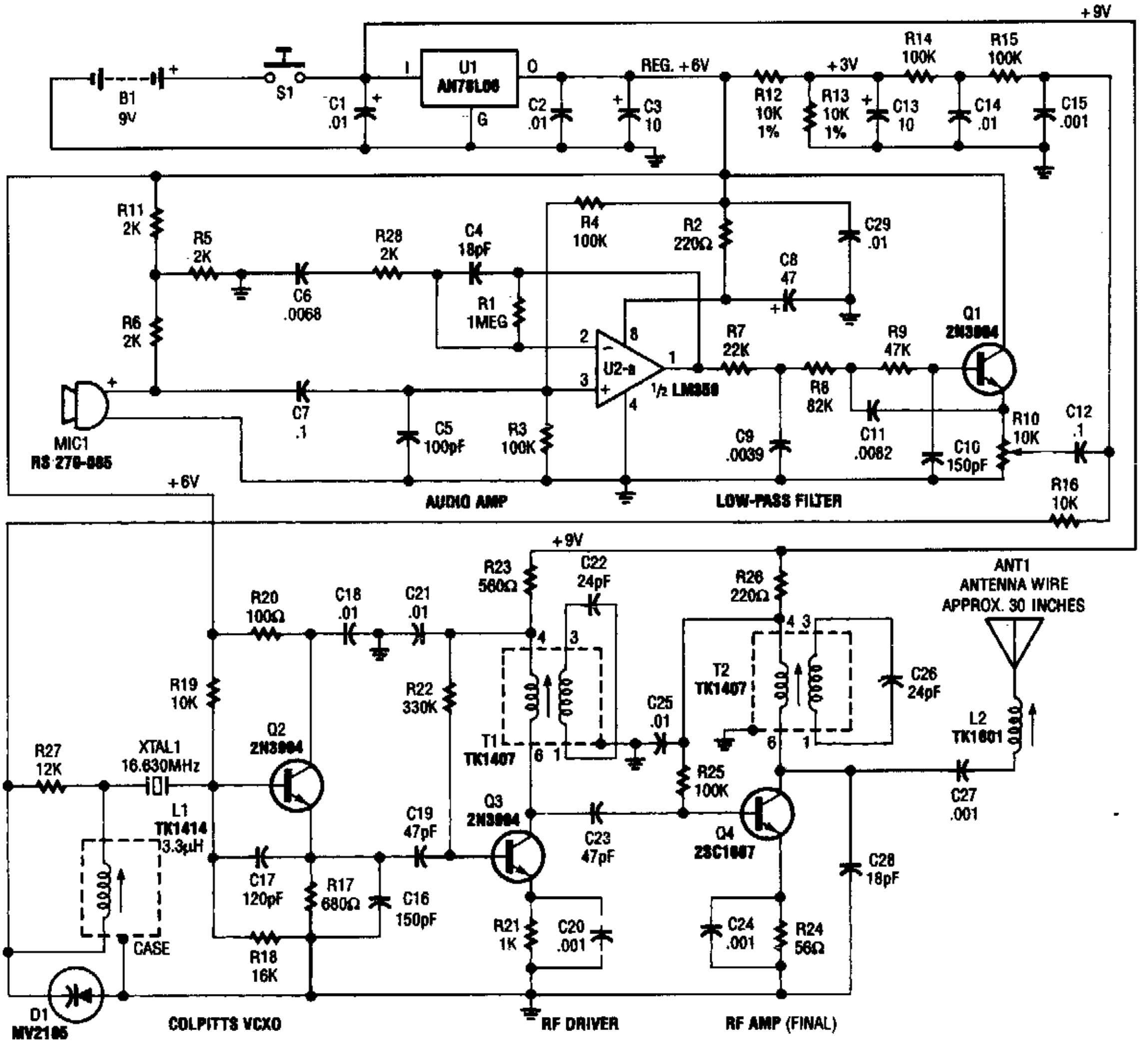
TRACKING TRANSMITTER



1993 ELECTRONICS HOBBYIST HANDBOOK ***FIG. 106-7***

This tracking transmitter consists of four distinct subassemblies; a free-running multivibrator, a transmit switch, an audio-tone generator, and an FM transmitter. The multivibrator (which produces a pulse width with a pulse separation of 1500 ms) is built around Q1 and Q2. The multivibrator output is coupled through R5 to the base of Q3, whose emitter feeds Q4, which controls the circuit's transmitter section.

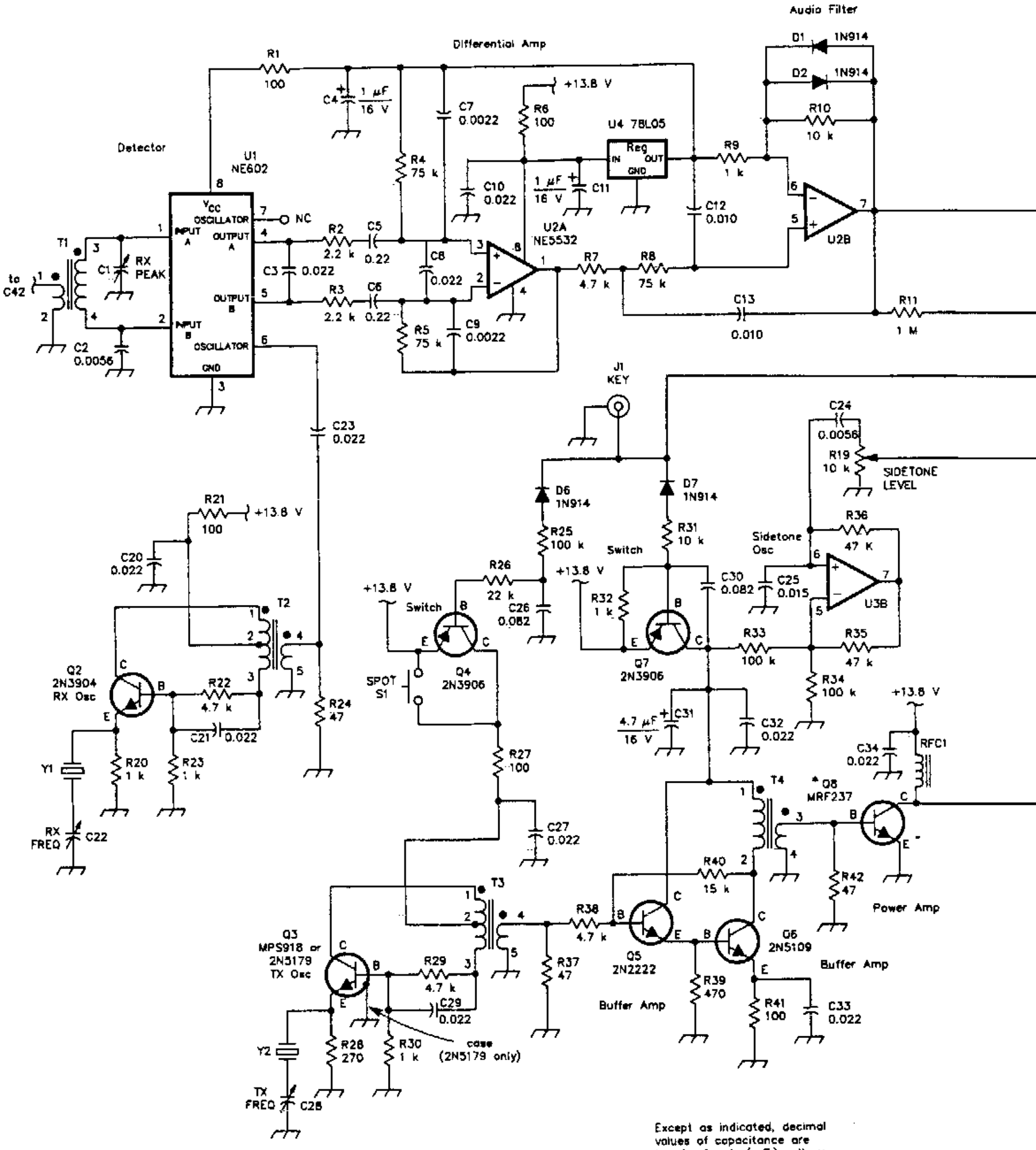
49-MHz FM TRANSMITTER



1993 ELECTRONICS HOBBYIST HANDBOOK ***FIG. 106-8***

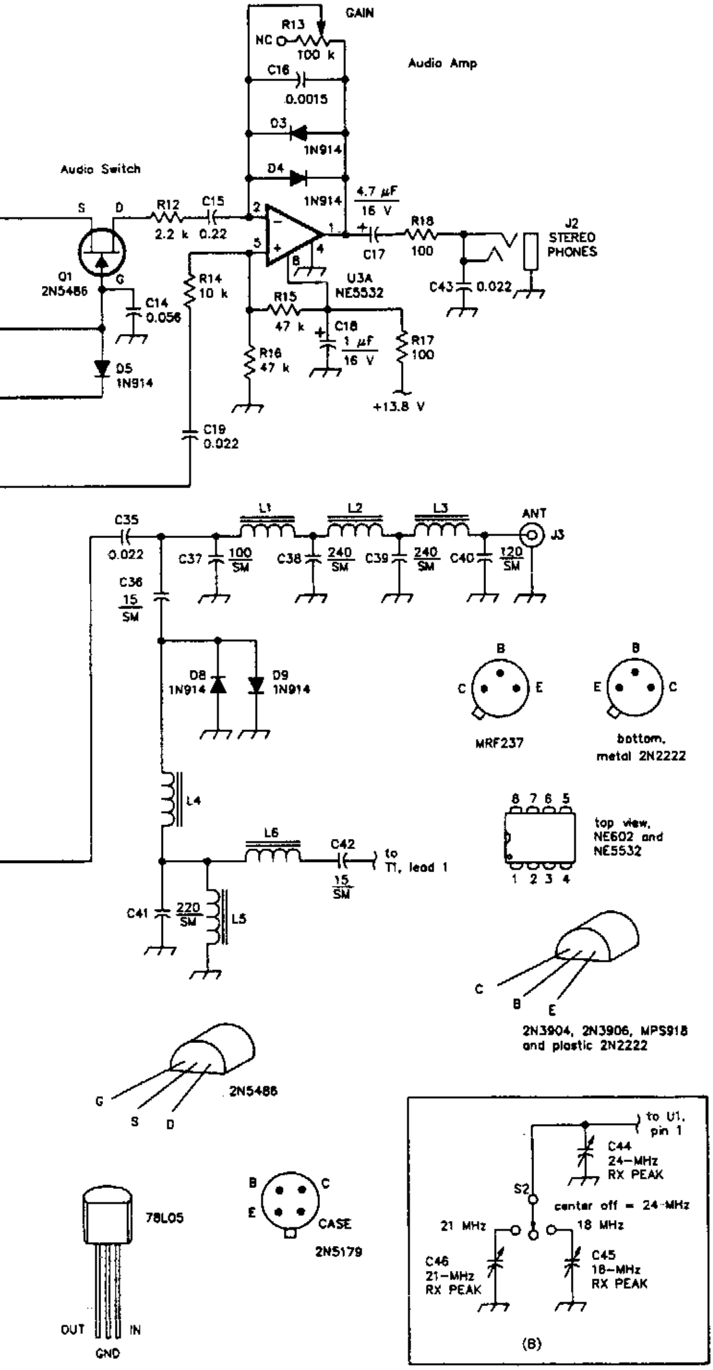
This 49-MHz FM transmitter consists of an audio amplifier, a low-pass filter, three RF stages, and a regulated-dc power supply. The output is about 16 mW into a 50-Ω load. This transmitter can be used in many 49-MHz applications, such as in a baby monitor, cordless telephone, or in conjunction with a scanner as a one-way voice link.

QRP TRANSCEIVER FOR 18, 21, AND 24 MHz



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000

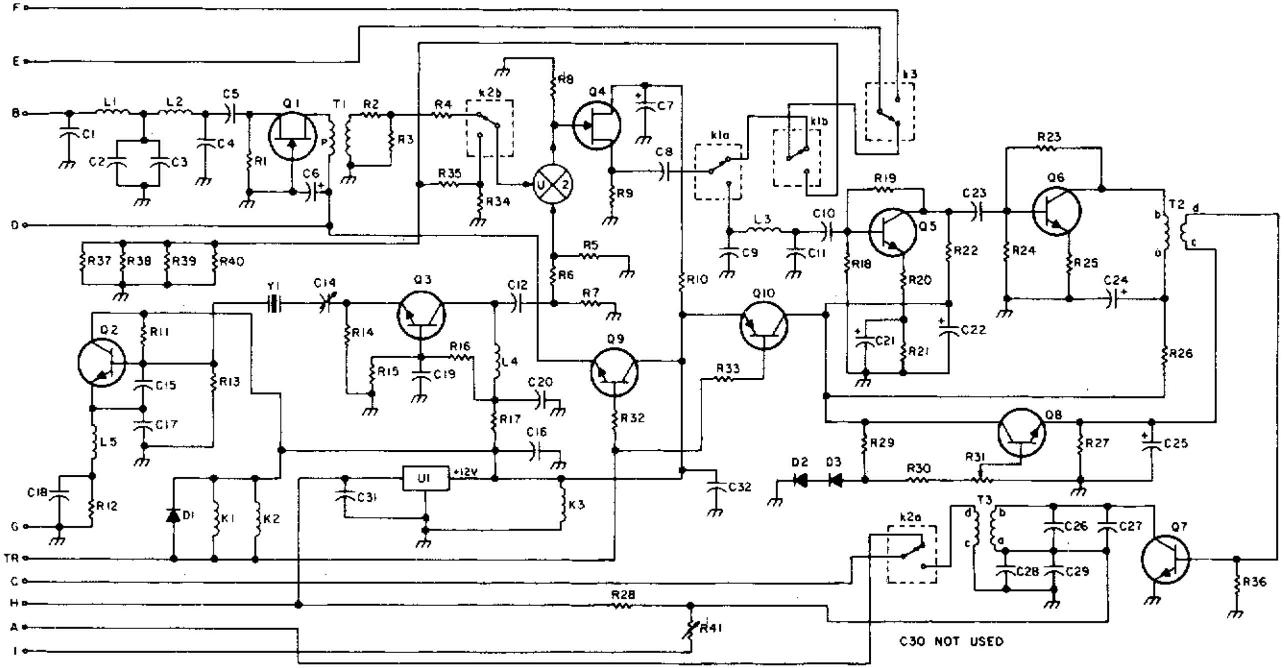
*Heat sink required; see text
● = phasing
SM = silver mica



This CW transceiver has 1.25 to 4 W RF output, a direct-conversion receiver, full break-in, and SW sidetone generation. The power supply is 13.8 V, which makes this transceiver suitable for mobile or portable operation.

FIG. 106-9

1750-METER TRANSVERTER

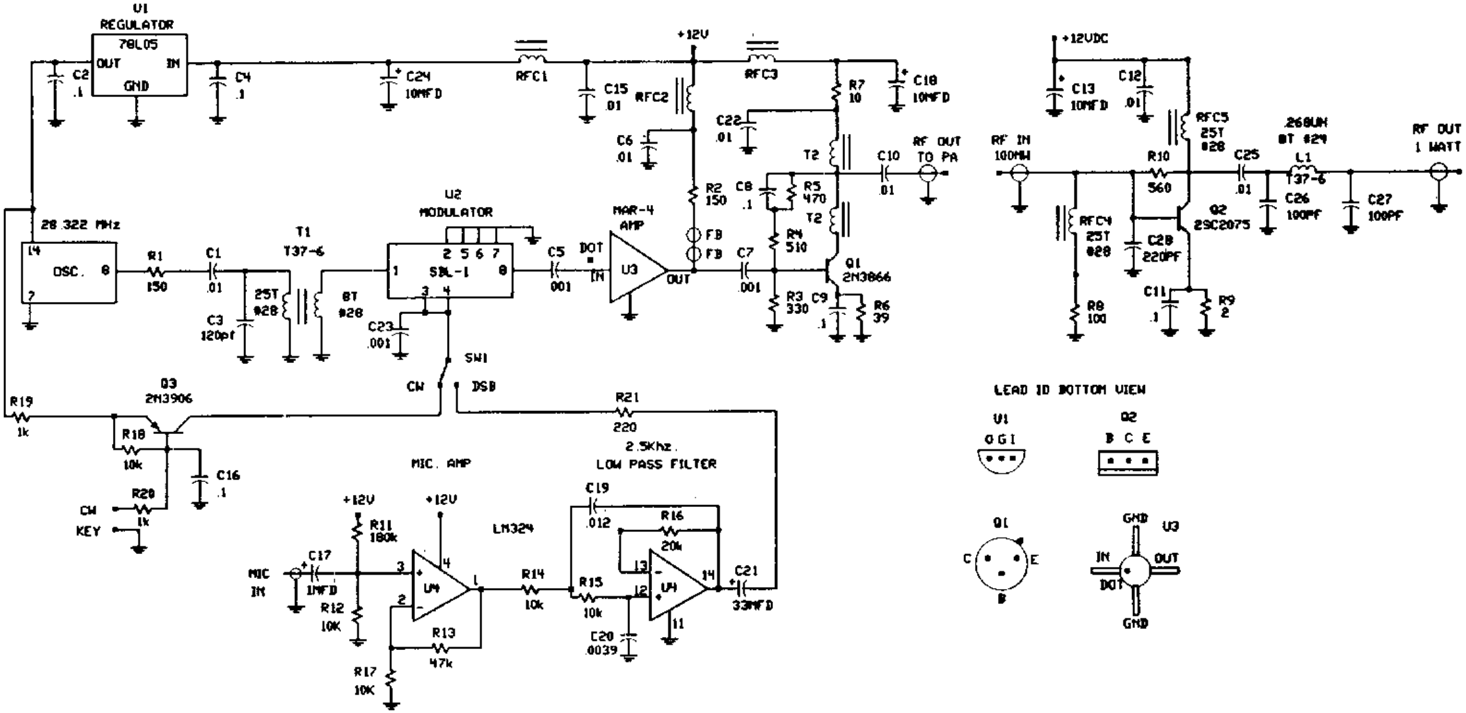


73 AMATEUR RADIO TODAY

FIG. 106-10

This circuit was described in a recent edition of an amateur radio magazine. It allows operation in the 160- to 190-kHz band with up to 1 W (license free) in any mode (CW/SSB/FM, etc.). It consists of a receiving converter for 5 kHz to 450 kHz and a transmitting converter to convert the 3.66- to 3.69-MHz (80 meter) range to 160 to 190 kHz. A 12- to 24-V power supply can be used.

10-METER DSB TRANSMITTER

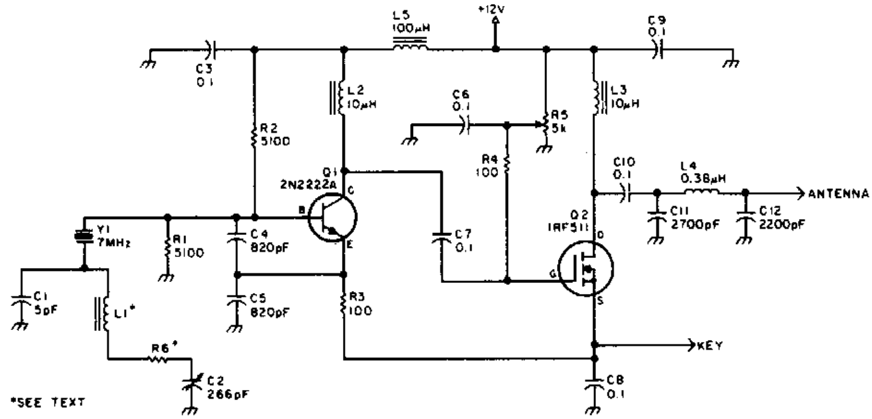


73 AMATEUR RADIO TODAY

FIG. 106-11

A DSB transmitter is much cheaper to build than an SSB transmitter because no filter or phasing networks are required. This circuit produces up to 1-W output on the 10-meter band. The frequency 28.322 MHz is used, which is a commonly available clock frequency crystal. CW operation is also provided. A doubly balanced mixer assembly is used as a modulator and CW keyer.

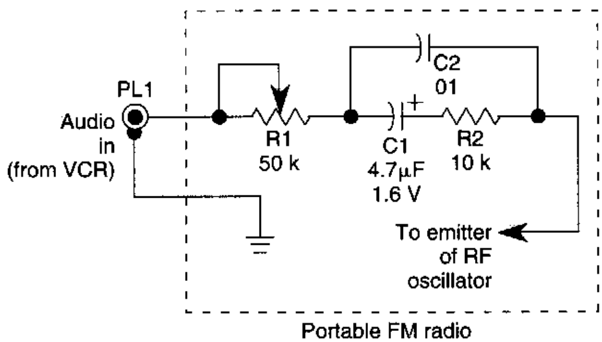
LOW-POWER 40-METER CW TRANSMITTER



73 AMATEUR RADIO TODAY ***FIG. 106-12***

This CW transmitter has an output of up to 3 W. By using 24 V on Q2, up to 10 W output can be obtained. If a 24-V supply is used, Q1 must not see more than 12 V. Connect 12 V between junctions C3, R2 and L2, and remove L5. L1 should be a low-*Q* 18- to 20-μH inductor. R6 can be used (up to 47 Ω) to reduce the *Q* further.

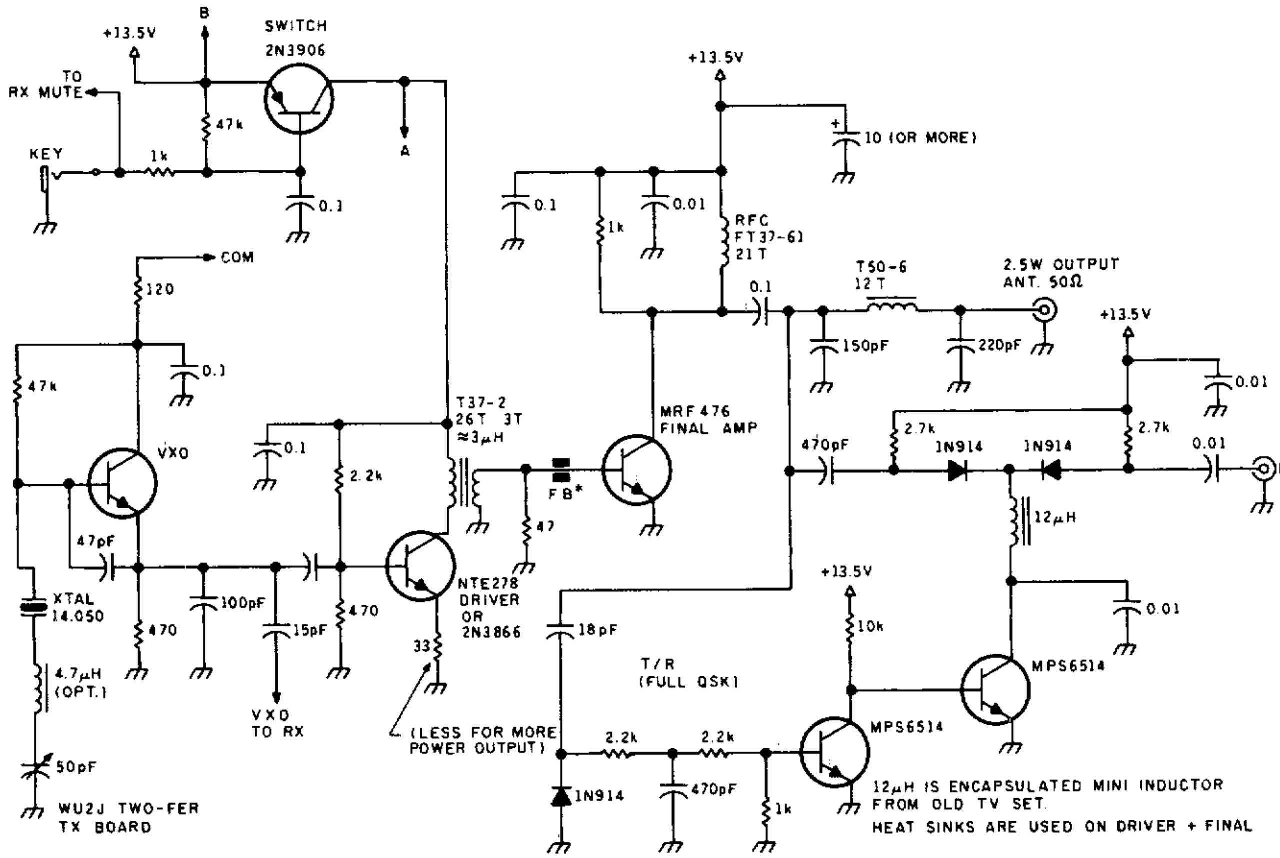
FM RADIO TRANSMITTER



R-E EXPERIMENTERS HANDBOOK ***FIG. 106-13***

An FM radio generates an interference signal that can be picked up on another FM radio tuned 10.7 MHz above the first one. The 50-kΩ potentiometer adjusts the modulation level to maximum without distortion. The RC network improves the fidelity of the transmitted signal and provides dc isolation. The component values shown are provided as a starting point. They can vary somewhat for different radios. Note that if you can't get the signal at 10.7 MHz above the frequency setting of the first radio, try tuning at 10.7 MHz below. Also, note that both tuned frequencies must be unused. Otherwise, you will hear your audio on top of the audio that is already there. You might have to play with both frequencies until you find two blank spots that are 10.7 MHz apart.

LOW-POWER 20-METER CW TRANSMITTER



73 AMATEUR RADIO TODAY ***FIG. 106-14***

The transmitter has a VXO circuit to drive an amplifier that is keyed. The keyed amplifier drives an MRF 476 final amplifier, which delivers about 2-W output. A solid-state T-R switch is included for the receiver. The parts values shown are for the 20-meter band.

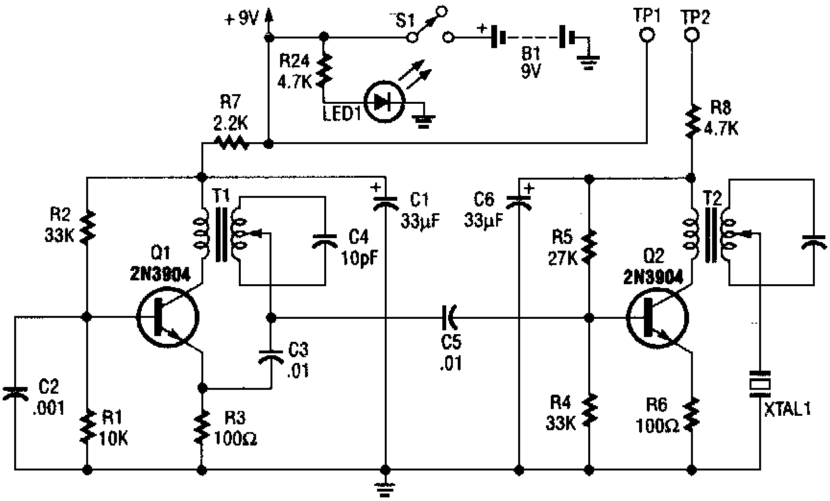
107

Ultrasonic Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

Doppler Ultrasound Transmitter
Doppler Ultrasound Receiver
Ultrasonic Cleaner

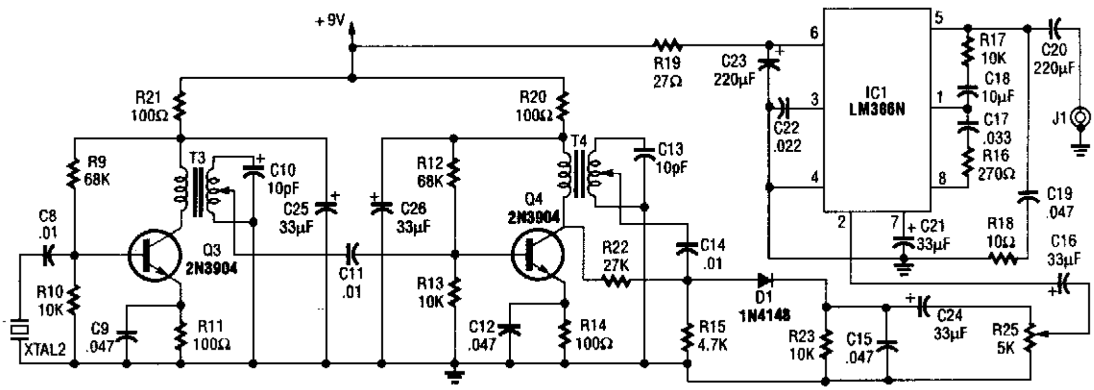
DOPPLER ULTRASOUND TRANSMITTER



RADIO-ELECTRONICS ***FIG. 107-1***

The 2.25-MHz oscillator Q1 drives amplifier Q2 and XTAL1, an ultrasonic transducer. The transducer is a lead zirconate-titanate type. Taps on T1 and T2 provide low-impedance drive points.

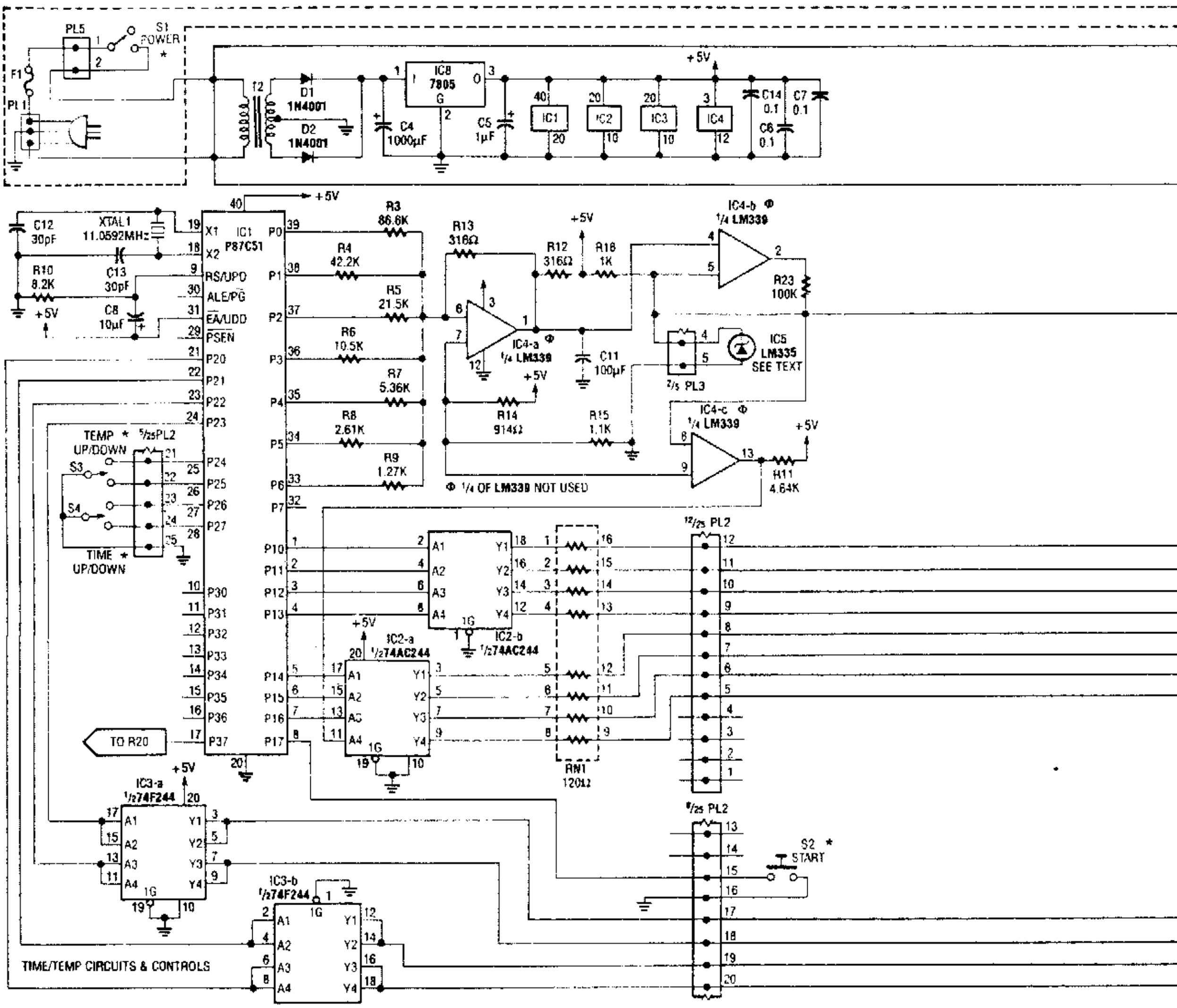
DOPPLER ULTRASOUND RECEIVER



RADIO-ELECTRONICS ***FIG. 107-2***

XTAL1 drives amplifier Q3/Q4, which is tuned to 2.25 MHz. The detected signal is fed to audio amplifier IC1. A 9-V supply is used. The circuit operates at 2.25 MHz and is designed to be used with an ultrasonic sound transmitter at this frequency.

ULTRASONIC CLEANER



ELECTRONICS NOW

An ultrasonic cleaner is useful to clean certain items. This circuit uses a microcontroller to control timing and give a digital readout, but only the basic oscillator can be used, if desired. RES1, RES2 are piezoelectric transducers driven by power oscillator Q1. Q1 is powered by a bridge rectifier-capacitor input filter that operates directly off the ac line. The frequency is 40 to 60 kHz.

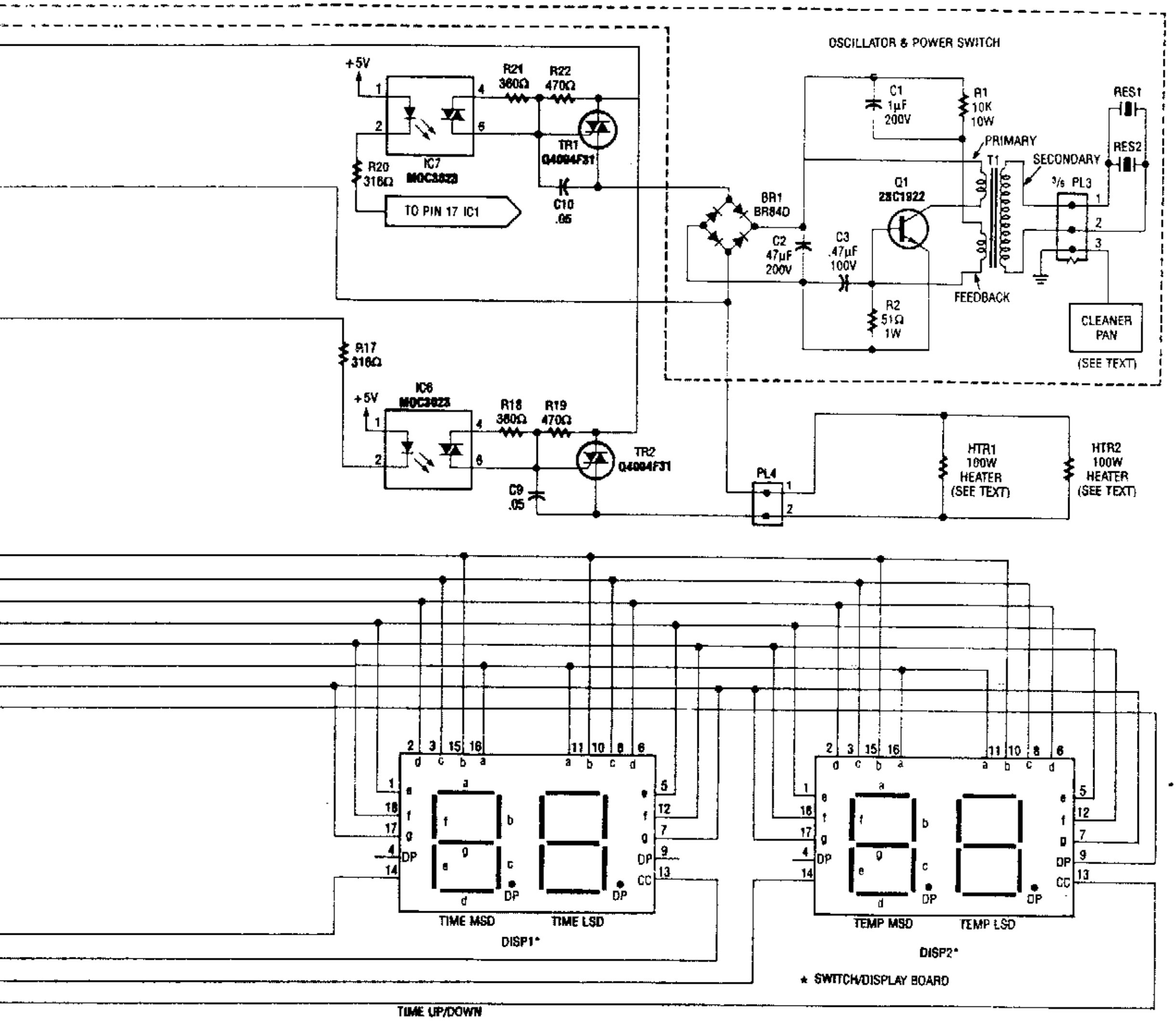

OSCILLATOR & POWER SWITCH
+5V
R21 360Ω
R22 470Ω
IC7 MOC3023
TR1 Q4004F31
R20 316Ω
TO PIN 17 IC1
C10 .05
C1 1µF 200V
R1 10K 10W
RES1
RES2
PRIMARY
SECONDARY
T1
PL3
Q1 2SC1922
BR1 BR84D
C2 47µF 200V
C3 .47µF 100V
R2 51Ω 1W
FEEDBACK
CLEANER PAN
(SEE TEXT)
R17 316Ω
IC6 MOC3023
R18 360Ω
R19 470Ω
TR2 Q4004F31
C9 .05
PL4
HTR1 100W HEATER (SEE TEXT)
HTR2 100W HEATER (SEE TEXT)
TIME MSD
TIME LSD
DISP1*
TEMP MSD
TEMP LSD
DISP2*
* SWITCH/DISPLAY BOARD
TIME UP/DOWN

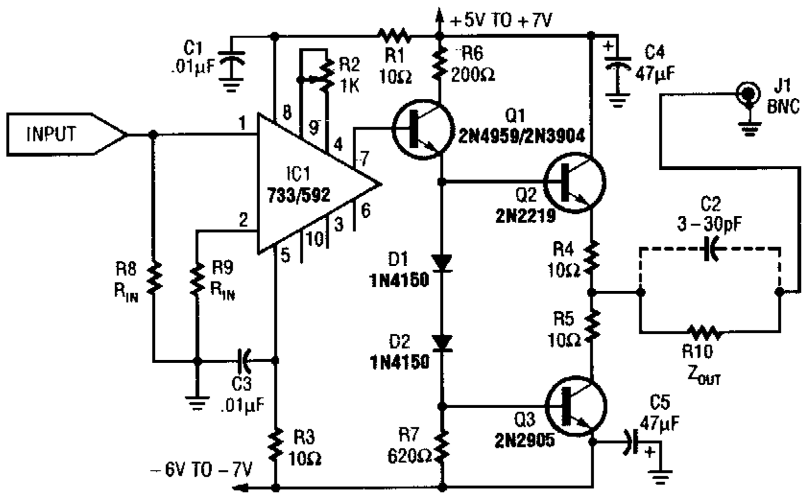
FIG. 107-3

108

Video Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

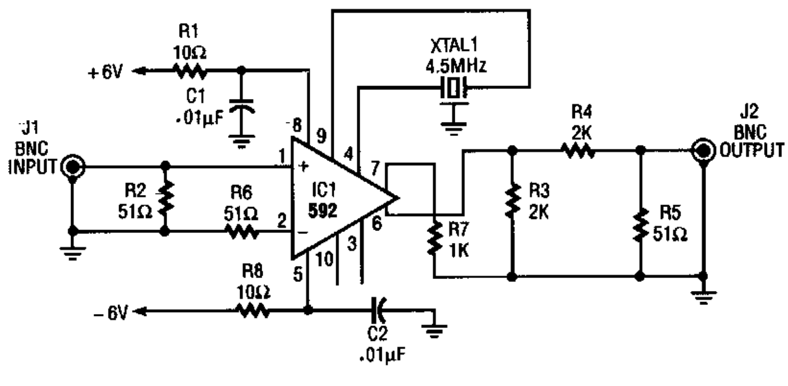
GENERAL-PURPOSE OUTPUT AMPLIFIER



RADIO-ELECTRONICS *FIG. 108-1*

This general-purpose amplifier has a bandwidth of approximately 20 MHz and it uses an LM733/NE592 video amp IC. This circuit can be used as a line driver or as a LAN line driver.

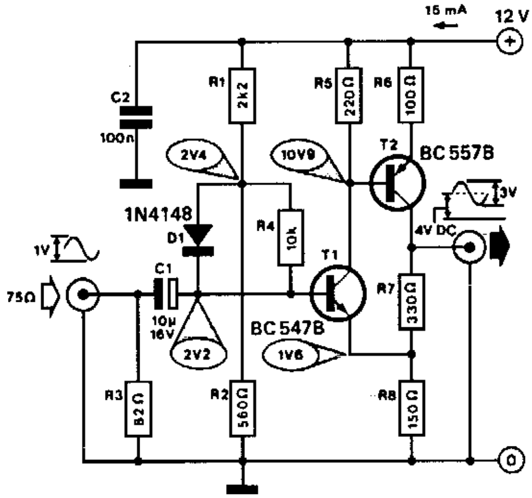
4.5-MHz SOUND IF AMPLIFIER



RADIO-ELECTRONICS *FIG. 108-2*

An NE592 is used as a 4.5-MHz amplifier sound subcarrier in video applications. XTAL1 is a 4.5-MHz crystal or ceramic resonator.

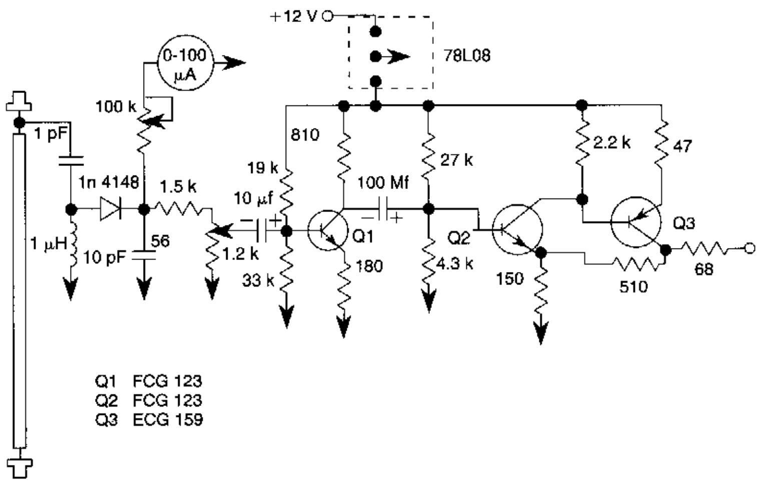
SIMPLE VIDEO AMPLIFIER



303 CIRCUITS *FIG. 108-3*

Useful for interfacing B/W TV sets with a camera or computer, this amplifier has a bandwidth of ≥10 MHz and a gain of 3X.

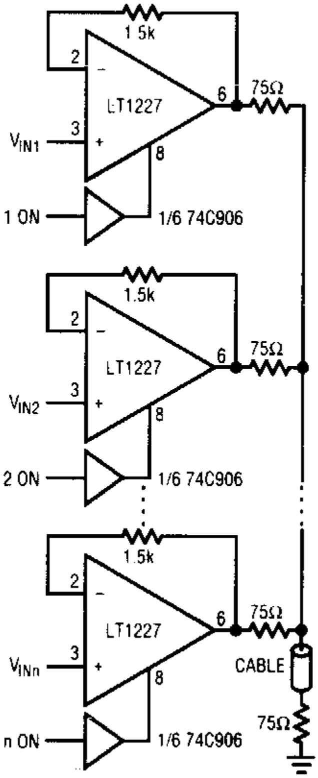
ATV VIDEO SAMPLER CIRCUIT



SPEC-COM *FIG. 108-4*

This unit picks up your ATV signal by sampling the transmission line with negligible insertion loss. It uses 2 "N" connectors for input and output connections. A BNC connector is used on the video output. The detected output is connected to your monitor and scope so that you can accurately adjust your transmitter for proper video and synch levels. Two different models are provided. Both have relative power output meters, but one has greater accuracy. There are two PC controls, one for video level and the other for power output.

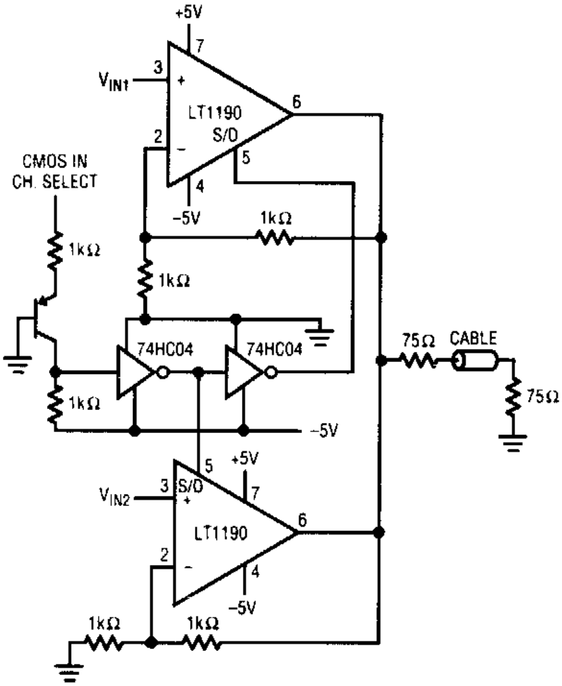
MULTIPLE-INPUT VIDEO MULTIPLEX CABLE DRIVER



LINEAR TECHNOLOGY ***FIG. 108-5***

Using a Linear Technology LT1227, the multiplex video amp uses logic levels to turn on and off selected inputs.

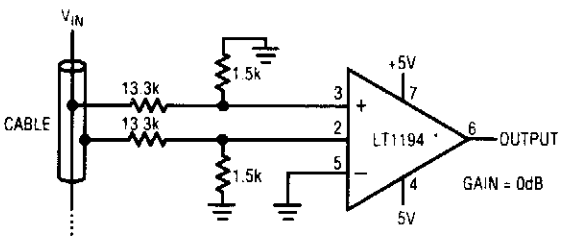
TWO-INPUT VIDEO MULTIPLEX CABLE DRIVER



LINEAR TECHNOLOGY ***FIG. 108-6***

CMOS logic levels select one of two video inputs with this circuit. The op amps are Linear Technology LT1190s.

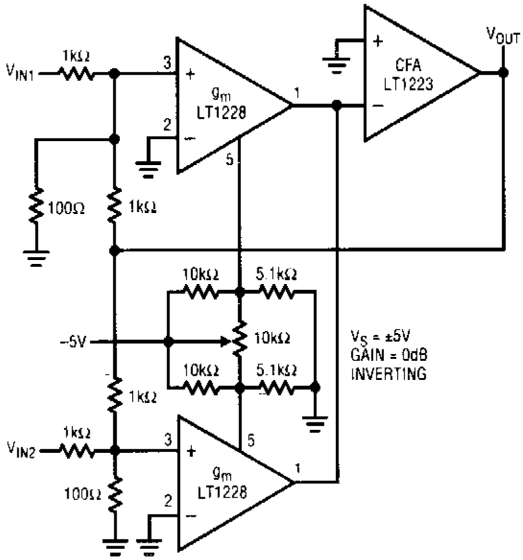
DIFFERENTIAL VIDEO LOOP-THROUGH AMPLIFIER



LINEAR TECHNOLOGY ***FIG. 108-7***

An LT1194 is used as a differential amplifier for video applications, where low cable loading is needed.

VIDEO FADER

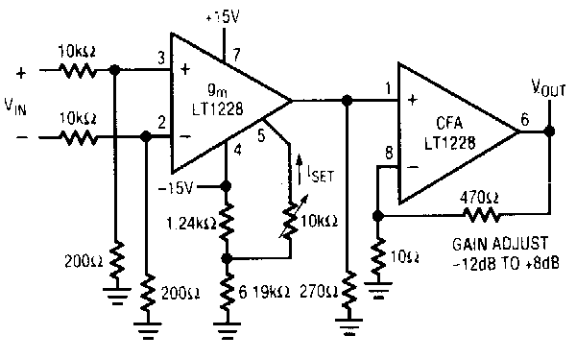


LINEAR TECHNOLOGY

FIG. 108-8

Using two LT1228 transconductance amplifiers in front of a current feedback amplifier forms a video fader. The ratio of the set currents into pin 5 determines the ratio of the inputs at the output.

ELECTRONICALLY CONTROLLED VARIABLE-GAIN VIDEO LOOP-THROUGH AMPLIFIER

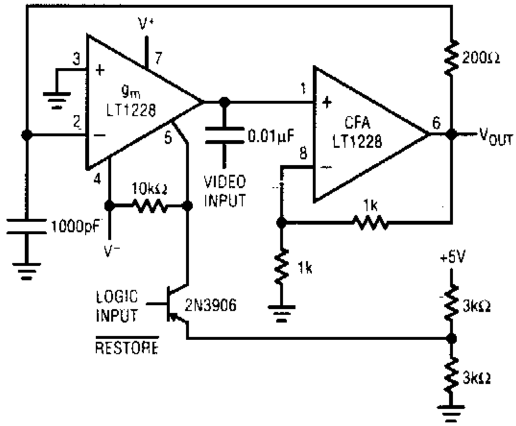


LINEAR TECHNOLOGY

FIG. 108-9

An LT1228 transconductance amplifier is used in this application. The gain is adjustable from −12 to +8 dB.

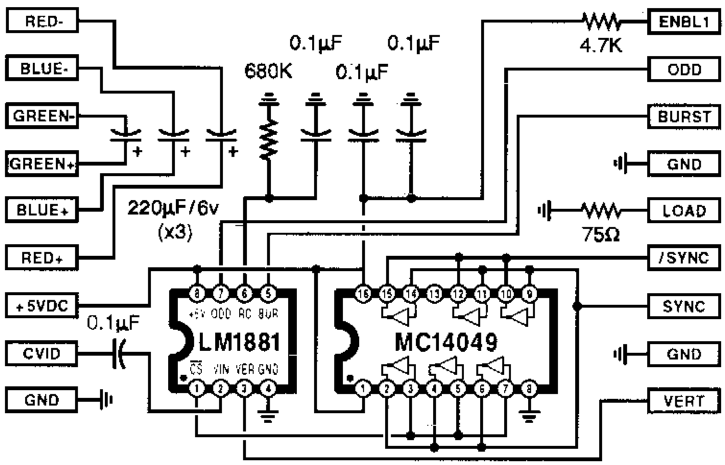
VIDEO dc RESTORE CIRCUIT



LINEAR TECHNOLOGY **_FIG. 108-10_**

This circuit restores the black level of a monochrome composite video signal to 0 V at the beginning of every horizontal line. This circuit is also useful with CCD scanners to set the black level.

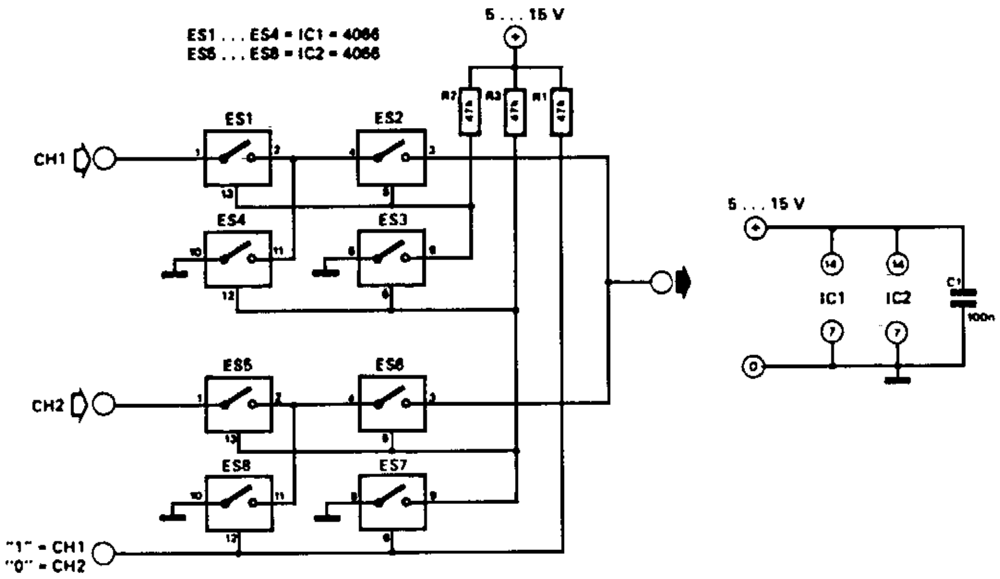
COMBINATION SYNC STRIPPER AND UNIVERSAL VIDEO INTERFACE



RADIO-ELECTRONICS **_FIG. 108-11_**

This combination sync stripper and universal video interface can solve a lot of problems for you, including Super-Nintendo-to-anything interfacing, video overlay and scope TV frame locking. Kits, fully tested units, and custom cable assemblies are available through Redmond Cable. This unit uses an LM1881 (NS) synch separator IC.

VIDEO SELECTOR

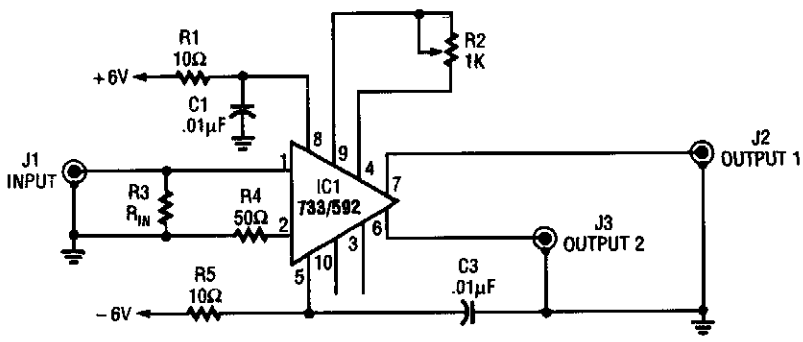


303 CIRCUITS

FIG. 108-12

This circuit selects one of two channels with a logic signal. The unused channel is shorted out, which minimizes crosstalk. The bandwidth at –3 dB is about 8 MHz. It is advisable to buffer this circuit because there is some loss in the switches when feeding a 75-Ω load.

VIDEO PREAMP

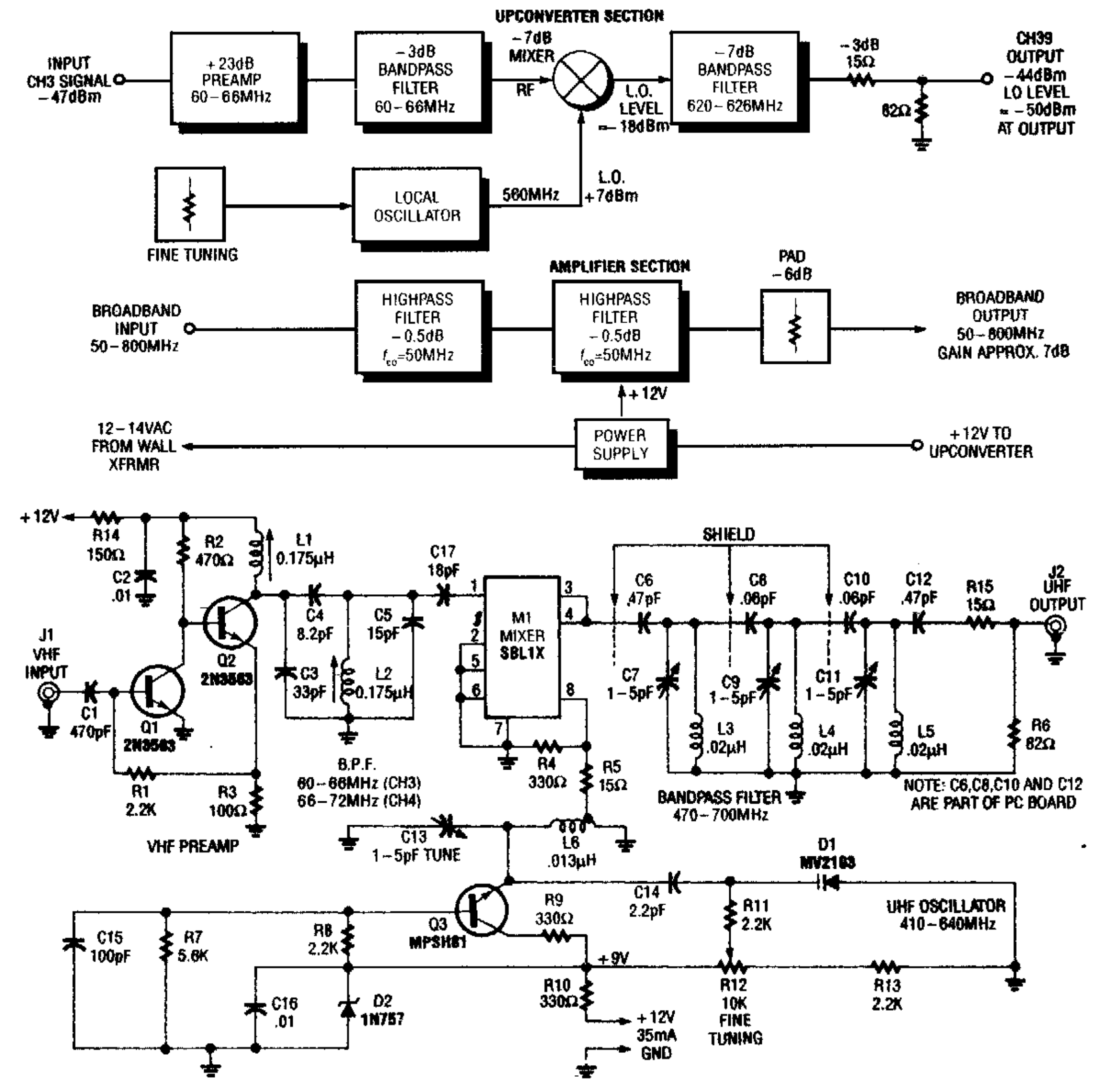


RADIO-ELECTRONICS

FIG. 108-13

An NE592 or LM733 is used as a general-purpose video amplifier in this schematic. J2 and J3 provide two anti-phase outputs. R2 is a gain control. The bandwidth is about 100 MHz.

VIDEO MASTER

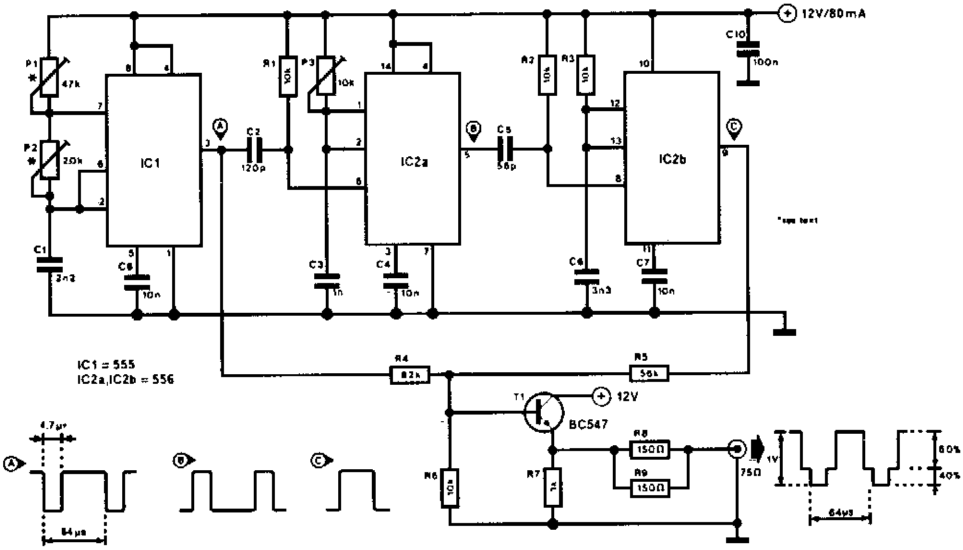


ELECTRONICS NOW *FIG. 108-14*

The video master consists of a series of converters that place all your video sources on unused UHF channels, which then combines them with normal TV channels (terrestrial or cable into one cable). That one cable can then feed several TV sets for whole-house coverage. The desired video source is selected with the TV set's tuner. All of the TV's remote-control features are retained.

A complete kit of parts is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

SIMPLE VIDEO LINE/BAR GENERATOR

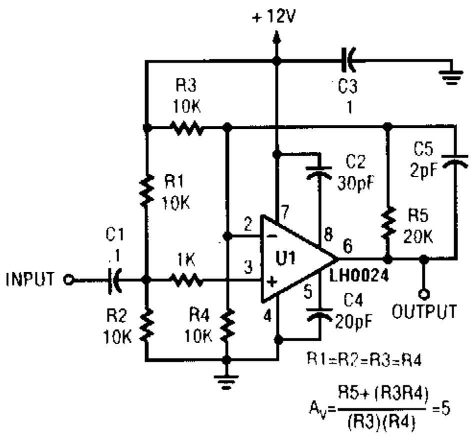


303 CIRCUITS

FIG. 108-15

A 555 and a dual 556 timer generate a rudimentary video signal, as shown in the schematic. The first timer generates 4.7-μs synch pulses operating in the astable mode with a 64-μs period. The second timer generates a delay pulse, which triggers the third timer to generate a bar. The second timer sets the bar position and the third sets the bar width.

VIDEO AMPLIFIER



POPULAR ELECTRONICS

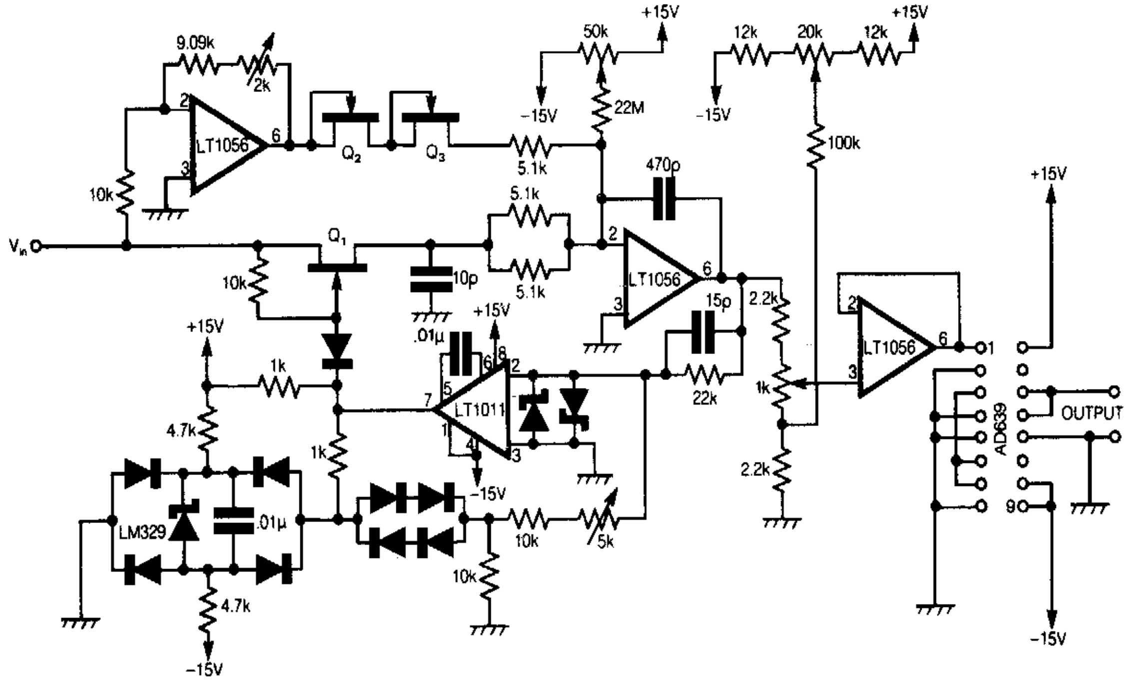
FIG. 108-16

109

Voltage-Controlled Oscillator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

SINUSOIDAL 3-Hz TO 300-kHz VCO



ELECTRONIC ENGINEERING ***FIG. 109-1***

This circuit uses Analog Devices' AD639 universal trigonometric function generator to convert a triangle waveform, the basic waveform of the VCO itself, into a very low-distortion sine wave.

By using the AD639 in its frequency tripler mode [2], the frequency range 3 Hz to 300 kHz is now covered. The circuit has been drawn here so that the oscillator loop, consisting of Q1, the integrator and the LT1011 comparator, is clearly shown.

When Q1 is off, the input amplifier, which is adjusted to have a gain of exactly −1, pulls a current V_{IN}/R, where R is 5.1 kΩ in series with two JFETs, and Q2 and Q3, out of the virtual earth of the integrator. The output of the integrator thus rises at a rate of V_{IN}/CR, where C = 470 pF. At a level that can be adjusted by the 5-kΩ potentiometer, the comparator flips and turns on Q1.

A current of exactly $2V_{IN}/R$, is now supplied to the virtual earth of the integrator because there are now two 5.1-kΩ resistors in parallel and only a single JFET in between the virtual earth and V_{in}. The integrator output now falls at a rate of V_{IN}/CR and the cycle repeats. Any offset in the current to the virtual earth of the integrator, due to circuit board leakage, etc., can be corrected by adjusting the 50-kΩ potentiometer. It follows that the symmetry of the triangle wave at the integrator output can be corrected by adjusting the 2-kΩ potentiometer, and the 50-kΩ potentiometer at VLF, and the frequency can be trimmed with the 5-kΩ potentiometer.

SINUSOIDAL 3-Hz TO 300-kHz VCO (*Cont.*)

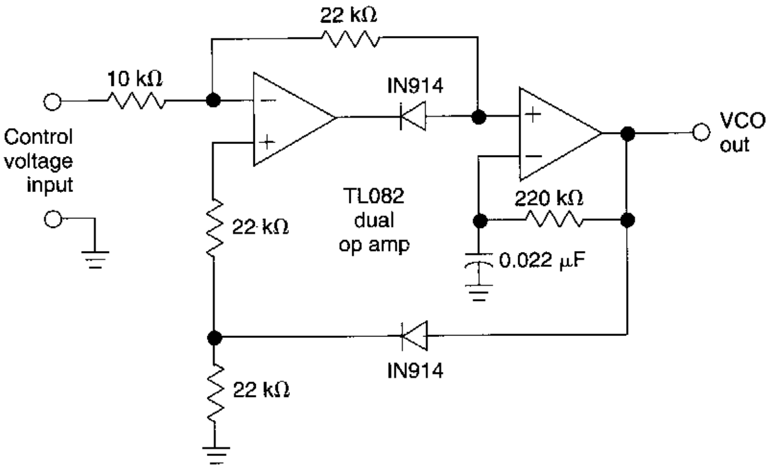
The 1-kΩ potentiometer variable is adjusted to give the input level to the AD639 needed to drive it over ±270° and so produce a sinusoidal output at three times the frequency of the triangle-wave input. Offset correction for the AD639 is made at the input to the voltage follower by means of the 20-kΩ potentiometer.

Once a symmetric triangle wave has been obtained by adjusting the 2-kΩ and 50-kΩ potentiometers, and the correct frequency of 100 kHz has been set for V_{IN} = 10 V, by adjusting the 5-kΩ potentiometer, the triple-frequency sine-wave output can be set up by adjustment of the 1-kΩ and 20-kΩ potentiometers.

This is best done by triggering the CRO from the triangle wave, and then viewing at least three complete cycles of output. Having adjusted for a clean-looking sine wave, the final adjustment of the 1-kΩ and 20-kΩ potentiometers should be made on a single sinusoidal cycle display, using internal trigger so that the three slightly different parts of the output cycle lie one upon the other and can be made to merge. Q1, Q2, and Q3 are 2N4391s, the two Schottky diodes are 5082–2810, and the other nine diodes are 1N914.

All device power supply pins should be decoupled with 0.33 µF. Resistors associated with the inputs of the devices should be 1% high-stability parts.

SIMPLE TL082 VCO

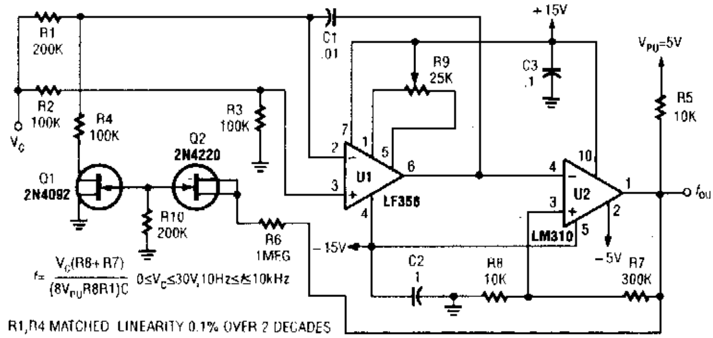


WILLIAM SHEETS

FIG. 109-2

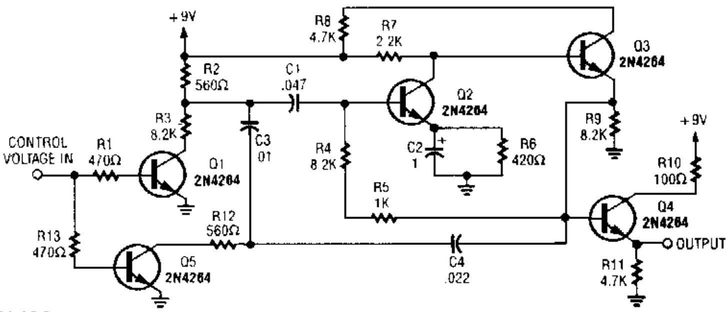
This circuit uses a dual operational amplifier (TL082) to form a voltage-controlled oscillator (VCO). With the component values shown, the output-frequency range is 100 Hz to 10 kHz when the input control voltage is between 0.05 and 10 V.

10-Hz TO 10-kHz 3-DECADE VCO



POPULAR ELECTRONICS *FIG. 109-3*

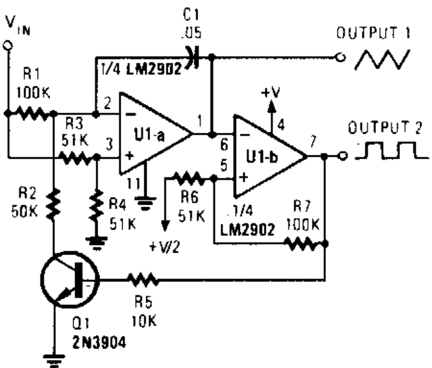
SINE-WAVE VCO



POPULAR ELECTRONICS *FIG. 109-4*

A dc control voltage varies the effective resistance in feedback network C4/C3/C1 and R12/R3. Q2/Q3 are the oscillator transistors.

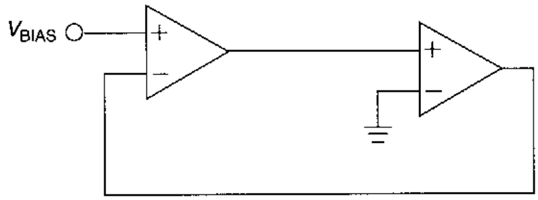
VCO I



This circuit gives both triangle- and square-wave outputs. The frequency range is determined by C1.

POPULAR ELECTRONICS *FIG. 109-5*

VCO II



WILLIAM SHEETS

FIG. 109-6

The output frequency of this simple low-cost active voltage-controlled oscillator circuit is based upon the inherent frequency dependent characteristics of our operational amplifier.

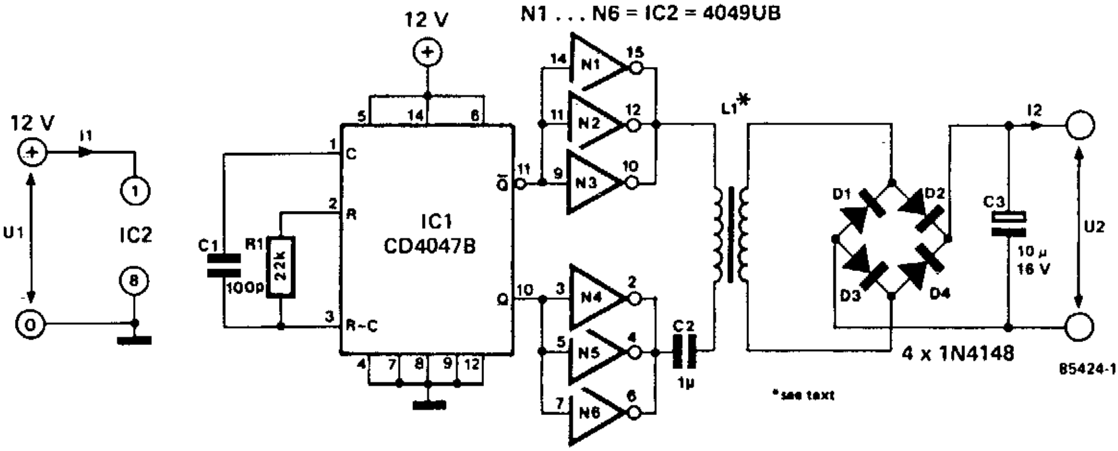
The oscillator circuit shown uses a TL082 op amp. When power is applied, the circuit generates a sinusoidal wave. The frequency of oscillation can be changed by varying the bias supply.

110

Voltage Converter/Inverter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

dc/dc CONVERTER

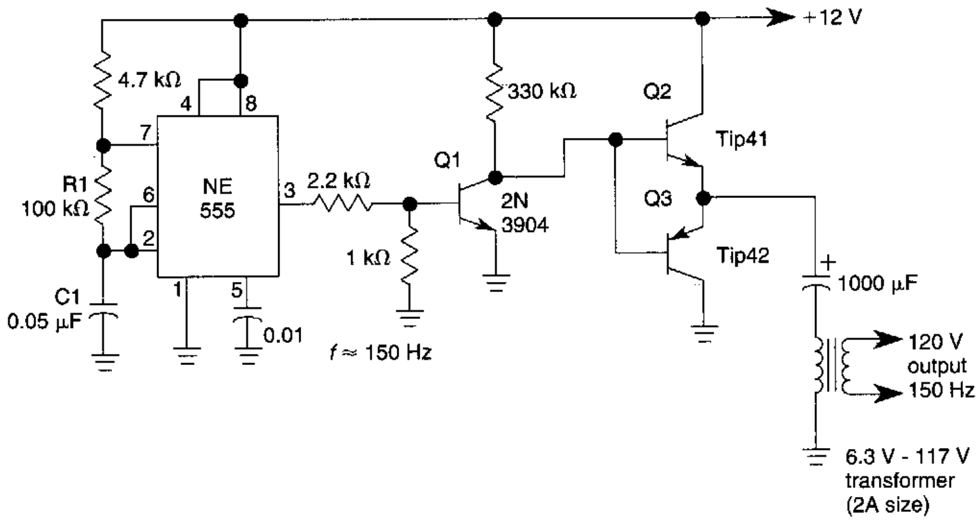


303 CIRCUITS

FIG. 110-1

This low-power converter will supply about 100 mW of dc to a load and it is useful to isolate or derive dc voltages. It operates at around 200 kHz. L1 is wound on a 22-mm diameter × 13-mm high pot core with #32 magnet wire. The primary is 80 turns and the secondary is 80 turns (for 12-V nominal output). The two windings should be insulated for the expected voltage difference between input and output in insulation applications.

SIMPLE dc/ac INVERTER



WILLIAM SHEETS

FIG. 110-2

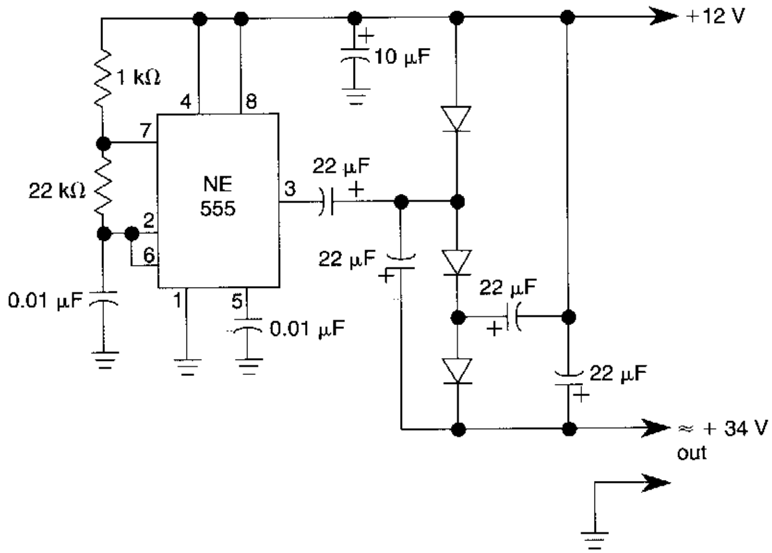
This dc-to-ac inverter is based on the popular 555. A 555 oscillator circuit drives a buffer amplifier consisting of Q1, Q2, and Q3. The circuit operates at 150 to 160 Hz. T1 can be a 6.3-V or 12.6-V filament transformer as applicable. The frequency can be changed by changing the values of R1 and/or C1.

111

Voltage Multiplier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

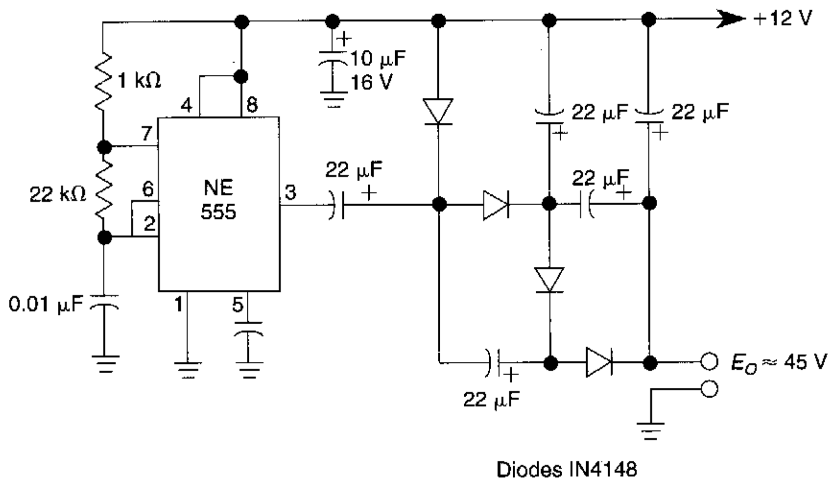
LOW-POWER dc TRIPLER



WILLIAM SHEETS *FIG. 111-1*

This dc voltage-tripler circuit based on the 555 can produce a dc output voltage equal to approximately 3× the dc supply voltage.

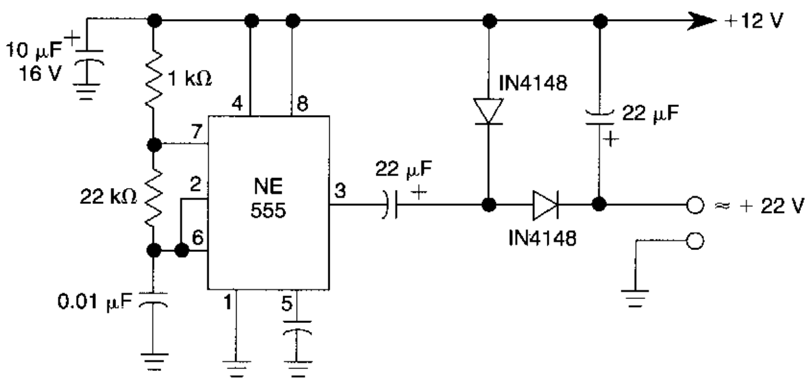
LOW-POWER dc QUADRUPLER



WILLIAM SHEETS *FIG. 111-2*

This dc voltage-quadrupler circuit based on the 555 can produce a dc output voltage equal to approximately 4× the dc supply voltage.

LOW-POWER dc DOUBLER



WILLIAM SHEETS

FIG. 111-3

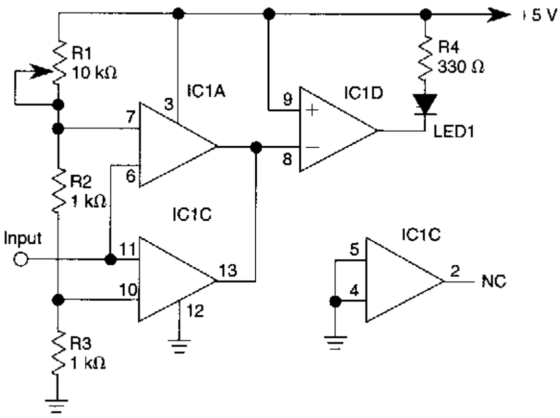
This dc voltage-doubler circuit based on the 555 can produce a dc output voltage equal to approximately 2× the dc supply voltage.

112

Window Comparator and Discriminator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 675. The figure number in the box of each circuit correlates to the entry in the Sources section.

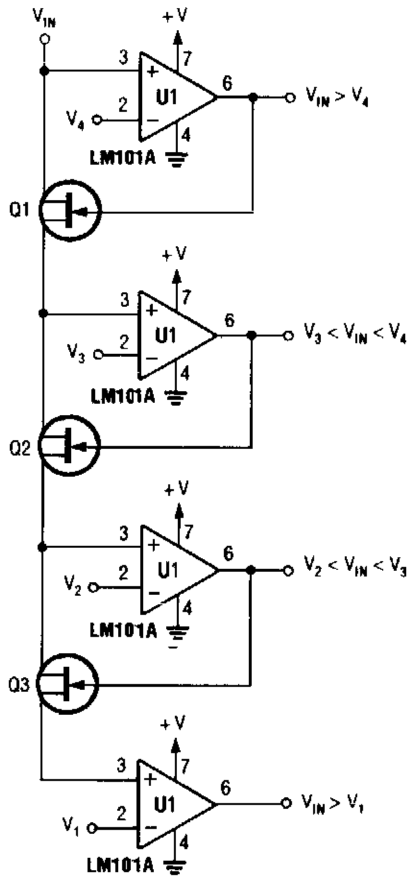
WINDOW COMPARATOR



WILLIAM SHEETS *FIG. 112-1*

IC1-c functions as a noninverting comparator, and IC1-a operates as an inverting comparator. Potentiometer R1 and fixed resistors R2 and R3 form a divider chain that delivers slightly different voltages to the two comparators. These voltages define the upper and lower limits of the circuit's switching "window," which can be changed easily by varying R2 and R3. The LED glows only when the input voltage falls within the window region.

MULTIPLE-APERTURE WINDOW DISCRIMINATOR



POPULAR ELECTRONICS *FIG. 112-2*

V1 through V4 are reference voltages that are derived from separate sources or from a common voltage divider.

Sources

Chapter 1

Fig. 1-1. Reprinted with permission from Popular Electronics, 1/92, p. 80. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-2. Reprinted with permission from Popular Electronics, 1/92, p. 80. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-3. Reprinted with permission from Popular Electronics, 1/92, p. 79. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-4. Reprinted with permission from Popular Electronics, 1/92, p. 79. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-5. Reprinted with permission from Popular Electronics, 1/92, p. 79. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-6. Reprinted with permission from Popular Electronics, 2/92, pp. 65-66. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-7. Reprinted with permission from Popular Electronics, 11/93, p. 53. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 1-8. Reprinted with permission from PE Hobbyist Handbook, 1992, pp. 93-94. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-9. William Sheets.

Fig. 1-10. Reprinted with permission from Popular Electronics, 2/92, pp. 70-71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-11. Reprinted with permission from Popular Electronics, 12/92, p. 68. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-12. William Sheets.

Fig. 1-13. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 31-32. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 1-14. Reprinted with permission from Popular Electronics, 8/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-15. Reprinted with permission from Popular Electronics, 2/92, p. 66. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-16. William Sheets.

Fig. 1-17. Reprinted with permission from Electronics Now, 7/92, p. 66. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 1-18. William Sheets.

Fig. 1-19. William Sheets.

Fig. 1-20. William Sheets.

Fig. 1-21. William Sheets.

Fig. 1-22. William Sheets.

Fig. 1-23. William Sheets.

Fig. 1-24. Reprinted with permission from Popular Electronics, 3/93, p. 42. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 1-25. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 19-20. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 1-26. Reprinted with permission from Popular Electronics, 1/92, p. 78. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 2

Fig. 2-1. Reprinted with permission from Popular Electronics, Fact Card 255, (c) Copyright Gernsback Publications, Inc.

Fig. 2-2. Reprinted with permission from Popular Electronics, Fact Card 254, (c) Copyright Gernsback Publications, Inc.

Fig. 2-3. Reprinted with permission from Popular Electronics, Fact Card 254, (c) Copyright Gernsback Publications, Inc.

Fig. 2-4. Reprinted with permission from Popular Electronics, Fact Card 253, (c) Copyright Gernsback Publications, Inc.

Fig. 2-5. Reprinted with permission from Popular Electronics, Fact Card 254, (c) Copyright Gernsback Publications, Inc.

Fig. 2-6. Reprinted with permission from Popular Electronics, Fact Card 254, (c) Copyright Gernsback Publications, Inc.

Fig. 2-7. Reprinted with permission from Popular Electronics, Fact Card 253, (c) Copyright Gernsback Publications, Inc.

Fig. 2-8. William Sheets.

Fig. 2-9. Reprinted with permission from Popular Electronics, Fact Card 253, (c) Copyright Gernsback Publications, Inc.

Fig. 2-10. William Sheets.

Fig. 2-11. William Sheets.

Fig. 2-12. William Sheets.

Fig. 2-13. William Sheets.

Fig. 2-14. William Sheets.

Fig. 2-15. Reprinted with permission from Electronics Now, 7/92, p. 36. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 2-16. Reprinted with permission from Popular Electronics, Fact Card 255, (c) Copyright Gernsback Publications, Inc.

Fig. 2-17. William Sheets.

Fig. 2-18. William Sheets.

Fig. 2-19. Reprinted with permission from Popular Electronics, Fact Card 223, (c) Copyright Gernsback Publications, Inc.

Fig. 2-20. William Sheets.

Fig. 2-21. William Sheets.

Fig. 2-22. Reprinted with permission from Popular Electronics, Fact Card 264, (c) Copyright Gernsback Publications, Inc.

Fig. 2-23. Reprinted with permission from Radio-Electronics, 6/92, p. 59. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 2-24. Reprinted with permission from 73 Amateur Radio Today, 4/92, p. 71.

Fig. 2-25. William Sheets.

Fig. 2-26. Reprinted with permission from National Semiconductor, Linear Edge, Spring 1992.

Fig. 2-27. Reprinted with permission from Popular Electronics, Fact Card 206, (c) Copyright Gernsback Publications, Inc.

Fig. 2-28. Reprinted with permission from Popular Electronics, 9/93, p. 47. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 3

Fig. 3-1. Reprinted with permission from Electronic Design, 3/93, p. 67.

Fig. 3-2. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 42.

Chapter 4

Fig. 4-1. Reprinted with permission from Popular Electronics, 12/91, p. 63. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 4-2. Reprinted with permission from Elector Electronics USA, 10/92, p. 14.

Fig. 4-3. Reprinted with permission from 73 Amateur Radio Today, 5/90, p. 47.

Fig. 4-4. Reprinted with permission from Practical Wireless, 6/91, p. 36.

Fig. 4-5. Reprinted with permission from Elektor Electronics, 12/91, pp. 88-89.

Fig. 4-6. Reprinted with permission from Popular Electronics, 6/93, p. 55. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 4-7. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 34.

Fig. 4-8. Reprinted with permission from Electronics Hobbyists Handbook, 1993, p. 89.

Fig. 4-9. Reprinted with permission from 73 Amateur Radio Today, 10/92, p. 28.

Chapter 5

Fig. 5-1. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 65-66. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 5-2. Reprinted with permission from Electronics Now, 11/92, p. 42. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-3. Reprinted with permission from Electronics Now, 12/92, p. 14. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-4. Reprinted with permission from Silicon Chip.

Fig. 5-5. Reprinted with permission from Electronics Now, 11/92, p. 43. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-6. Reprinted with permission from Electronics Now, 11/92, p. 41. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-7. Reprinted with permission from 303 Circuits, p. 42.

Fig. 5-8. Reprinted with permission from Electronics Now, 11/92, p. 39. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-9. Reprinted with permission from Electronics Now, 11/92, p. 39. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-10. Reprinted with permission from R-E Experimenters Handbook, 1987, p. 74.

Fig. 5-11. Reprinted with permission from Popular Electronics, 1/92, p. 36. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-12. Reprinted with permission from Popular Electronics, Fact Card 243, (c) Copyright Gernsback Publications, Inc.

Fig. 5-13. Reprinted with permission from Radio-Electronics, 1/92, p. 35. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-14. Reprinted with permission from Popular Electronics, 4/92, p. 69. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 5-15. Reprinted with permission from Electronic Design, 3/89, p. 100.

Chapter 6

Fig. 6-1. Reprinted with permission from 303 Circuits, p. 22.

Fig. 6-2. Reprinted with permission from 303 Circuits, p. 10.

Fig. 6-3. William Sheets.

Fig. 6-4. Reprinted with permission from 303 Circuits, p. 40.

Fig. 6-5. Reprinted with permission from Popular Electronics, 6/92, p. 68. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 6-6. Reprinted with permission from 303 Circuits, pp. 12-13.

Fig. 6-7. William Sheets.

Fig. 6-8. Reprinted with permission from R-E Experimenters Handbook, 1987, p. 74.

Fig. 6-9. Reprinted with permission from Popular Electronics, Fact Card 267, (c) Copyright Gernsback Publications, Inc.

Fig. 6-10. Reprinted with permission from Popular Electronics, Fact Card 267, (c) Copyright Gernsback Publications, Inc.

Fig. 6-11. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 37.

Fig. 6-12. Reprinted with permission from Popular Electronics, Fact Card 263, (c) Copyright Gernsback Publications, Inc.

Chapter 7

Fig. 7-1. Reprinted with permission from 73 Amateur Radio Today, 8/92, p. 36.

Fig. 7-2. William Sheets.

Fig. 7-3. Reprinted with permission from RF Communications Handbook, 1989, pp. 2-14.

Chapter 8

Fig. 8-1. Reprinted with permission from Popular Electronics, 9/92, p. 33. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-2. Reprinted with permission from Popular Electronics, 3/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-3. William Sheets.

Fig. 8-4. Reprinted with permission from Electronics Now, 9/93, p. 63. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 8-5. William Sheets.

Fig. 8-6. Reprinted with permission from Silicon Chip, p. 27.

Fig. 8-7. Reprinted with permission from PE Hobbyist Handbook, 1990, pp. 86-87. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 8-8. Reprinted with permission from Electronics Now, 11/92, p. 59. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-9. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 22.

Fig. 8-10. Reprinted with permission from Popular Electronics, 10/93, p. 64. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 8-11. Reprinted with permission from PE Hobbyist Handbook, p. 73. (c) Copyright Gernsback Publications, Inc.

Fig. 8-12. Reprinted with permission from Popular Electronics, 3/93, p. 62. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 8-13. Reprinted with permission from Popular Electronics, 11/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 8-14. Reprinted with permission from Popular Electronics, 3/92, p. 73. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-15. Reprinted with permission from Popular Electronics, 3/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-16. Reprinted with permission from Popular Electronics, 3/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-17. Reprinted with permission from Popular Electronics, 3/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-18. Reprinted with permission from Popular Electronics, 12/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-19. William Sheets.

Fig. 8-20. Reprinted with permission from Radio-Electronics, 12/91, p. 75. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 8-21. Reprinted with permission from Radio-Electronics, 5/92, p. 82. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-22. Reprinted with permission from Popular Electronics, 4/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 8-23. William Sheets.

Fig. 8-24. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 158.

Chapter 9

Fig. 9-1. Reprinted with permission from Popular Electronics, 6/93, p. 76. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 9-2. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 122.

Fig. 9-3. Reprinted with permission from Popular Electronics, 4/92, p. 71 & 88. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 9-4. Reprinted with permission from National Semiconductor, Linear Edge, Summer 1992.

Fig. 9-5. Reprinted with permission from Radio-Electronics, 5/92, p. 12. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 10

Fig. 10-1. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 44-45. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 10-2. Reprinted with permission from Electronics Now, 7/92, pp. 57-62. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 10-3. Reprinted with permission from Popular Electronics, Fact Card 198. (c) Copyright Gernsback Publications, Inc.

Fig. 10-4. Reprinted with permission from 73 Amateur Radio Today, 5/92, p. 26.

Fig. 10-5. Reprinted with permission from Electronic Design, 7/93, p. 78.

Fig. 10-6. Reprinted with permission from Popular Electronics, Fact Card 198. (c) Copyright Gernsback Publications, Inc.

Fig. 10-7. Reprinted with permission from Popular Electronics, 11/91, p. 20. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 10-8. Reprinted with permission from Linear Technology, Design Note 60.

Fig. 10-9. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 63-64. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 10-10. Reprinted with permission from Elektor Electronics, 12/91, p. 72.

Fig. 10-11. Reprinted with permission from Elektor Electronics USA, 12/91, p. 36.

Chapter 11

Fig. 11-1. Reprinted with permission from Popular Electronics, 8/93, p. 79. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 11-2. Reprinted with permission from Popular Electronics, 6/93, p. 70. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 11-3. Reprinted with permission from Popular Electronics, 8/93, p. 79. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 11-4. Reprinted with permission from Popular Electronics, 8/93, p. 79. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 11-5. Reprinted with permission from National Semiconductor, Linear Edge, Issue #5.

Fig. 11-6. Reprinted with permission from Electronics Now, 9/92, p. 96. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 11-7. William Sheets.

Chapter 12

Fig. 12-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 58.

Fig. 12-2. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 59.

Chapter 13

Fig. 13-1. Reprinted with permission from PE Hobbyist Handbook, 1992, p. 49. (c) Copyright Gernsback Publications, Inc.

Chapter 14

Fig. 14-1. Reprinted with permission from Popular Electronics, 3/93, p. 44. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 14-2. Reprinted with permission from 73 Amateur Radio Today, 5/92, p. 20.

Fig. 14-3. Reprinted with permission from 73 Amateur Radio Today, 5/92, p. 18.

Fig. 14-4. Reprinted with permission from Popular Electronics, Fact Card 206, (c) Copyright Gernsback Publications, Inc.

Fig. 14-5. William Sheets.

Fig. 14-6. Reprinted with permission from Popular Electronics, 10/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 14-7. Reprinted with permission from Popular Electronics, Fact Card 206, (c) Copyright Gernsback Publications, Inc.

Chapter 15

Fig. 15-1. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 24-26. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 16

Fig. 16-1. Reprinted with permission from Popular Electronics, 3/92, p. 60. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 16-2. Reprinted with permission from Electronics Now, 7/92, p. 51. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 16-3. Reprinted with permission from Electronic Design, 3/93, pp. 67-68.

Chapter 17

Fig. 17-1. Reprinted with permission from Elektor Electronics, 12/91, pp. 78-79.

Fig. 17-2. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 84.

Fig. 17-3. Reprinted with permission from Electronic Design.

Fig. 17-4. Reprinted with permission from Popular Electronics, 11/91, p. 20. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 17-5. Reprinted with permission from Popular Electronics, 6/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 18

Fig. 18-1. Reprinted with permission from Popular Electronics, 12/91, p. 58. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 18-2. Reprinted with permission from PE Hobbyist Handbook, 1990, pp. 34-36. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 18-3. Reprinted with permission from Electronic Design, 5/92, p. 91.

Fig. 18-4. Reprinted with permission from Popular Electronics, 6/92, p. 57. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 18-5. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 92.

Fig. 18-6. Reprinted with permission from Radio-Electronics, 3/92, p. 50. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 18-7. Reprinted with permission from 303 Circuits, p. 197.

Fig. 18-8. Reprinted with permission from Radio-Electronics, 2/92, p. 42. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 18-9. Reprinted with permission from Popular Electronics, 12/91, p. 58. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 18-10. Reprinted with permission from National Semiconductor, Linear Edge, Issue #4, Summer 1992.

Fig. 18-11. Reprinted with permission from 73 Amateur Radio Today, 2/93, p. 28.

Fig. 18-12. William Sheets.

Fig. 18-13. Reprinted with permission from 73 Amateur Radio Today, 3/92, p. 24.

Fig. 18-14. Reprinted with permission from Popular Electronics, 7/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 18-15. Reprinted with permission from Electronics Now, 9/92, p. 79. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 18-16. Reprinted with permission from National Semiconductor, Linear Edge, Spring 1992.

Fig. 18-17. Reprinted with permission from Popular Electronics, Fact Card 259, (c) Copyright Gernsback Publications, Inc.

Fig. 18-18. Reprinted with permission from Popular Electronics, Fact Card 257, (c) Copyright Gernsback Publications, Inc.

Fig. 18-19. Reprinted with permission from Linear Technology Corporation, 1993, Design Note 69.

Chapter 19

Fig. 19-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 47.

Fig. 19-2. Reprinted with permission from PE Hobbyist Handbook, 1990, p. 101. (c) Copyright Gernsback Publications, Inc., 1990.

Chapter 20

Fig. 20-1. Reprinted with permission from RF Design, 5/92, p. 80.

Fig. 20-2. William Sheets.

Fig. 20-3. Reprinted with permission from Electronic Design, 11/92, p. 61.

Fig. 20-4. Reprinted with permission from Electronics Now, 12/92, p. 12. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 20-5. Reprinted with permission from Radio-Electronics, 2/92, p. 89. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 20-6. Reprinted with permission from 73 Amateur Radio Today, 5/92, p. 64.

Fig. 20-7. Reprinted with permission from 73 Amateur Radio Today, 8/92, p. 48.

Fig. 20-8. Reprinted with permission from 73 Amateur Radio Today, 1/92, p. 22.

Fig. 20-9. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 60.

Fig. 20-10. William Sheets.

Fig. 20-11. Reprinted with permission from Popular Electronics, Fact Card 229, (c) Copyright Gernsback Publications, Inc.

Chapter 21

Fig. 21-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 78.

Fig. 21-2. Reprinted with permission from National Semiconductor, Linear Edge, Issue #5.

Fig. 21-3. Reprinted with permission from Popular Electronics, Fact Card 257, (c) Copyright Gernsback Publications, Inc.

Fig. 21-4. Reprinted with permission from National Semiconductor, Linear Edge, Issue #5.

Chapter 22

Fig. 22-1. Reprinted with permission from Silicon Chip.

Fig. 22-2. Reprinted with permission from Electronics Now, 12/92, p. 49. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 23

Fig. 23-1. Reprinted with permission from Elektor Electronics, 12/91, p. 81.

Chapter 24

Fig. 24-1. Reprinted with permission from PE Hobbyist Handbook, 1992, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 24-2. Reprinted with permission from Electronics Now, 12/92, p. 61. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 24-3. Reprinted with permission from Electronics Now, 12/92, p. 65. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 24-4. Reprinted with permission from Electronic Design, 5/92, p. 95.

Fig. 24-5. Reprinted with permission from Linear Technology, Design Note 61.

Fig. 24-6. Reprinted with permission from Linear Technology, Design Note 61.

Fig. 24-7. Reprinted with permission from Linear Technology, Design Note 61.

Fig. 24-8. Reprinted with permission from Popular Electronics, Fact Card, 270, (c) Copyright Gernsback Publications, Inc.

Fig. 24-9. Reprinted with permission from Popular Electronics, Fact Card, 269, (c) Copyright Gernsback Publications, Inc.

Fig. 24-10. Reprinted with permission from Popular Electronics, Fact Card, 270, (c) Copyright Gernsback Publications, Inc.

Fig. 24-11. Reprinted with permission from 73 Amateur Radio Today, 8/89, p. 48.

Chapter 25

Fig. 25-1. Reprinted with permission from PE Hobbyist Handbook, 1992, pp. 63-64. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 25-2. Reprinted with permission from 303 Circuits, p. 14.

Fig. 25-3. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 58.

Fig. 25-4. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 57.

Fig. 25-5. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 57.

Fig. 25-6. Reprinted with permission from Elector Electronics USA, 12/91, p. 36.

Chapter 26

Fig. 26-1. Reprinted with permission from Radio-Electronics, 5/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-2. Reprinted with permission from Radio-Electronics, 5/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-3. Reprinted with permission from Electronic Design, 8/92, p. 70.
Fig. 26-4. Reprinted with permission from Radio-Electronics, 5/92, p. 69. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-5. Reprinted with permission from Radio-Electronics, 5/92, p. 69. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-6. Reprinted with permission from Radio-Electronics, 5/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-7. Reprinted with permission from Radio-Electronics, 5/92, p. 69. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-8. Reprinted with permission from Radio-Electronics, 5/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-9. Reprinted with permission from Radio-Electronics, 5/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-10. Reprinted with permission from Radio-Electronics, 5/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-11. Reprinted with permission from Radio-Electronics, 5/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 26-12. Reprinted with permission from Radio-Electronics, 5/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 27

Fig. 27-1. Reprinted with permission from Popular Electronics, 9/93, p. 42. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 27-2. Reprinted with permission from 303 Circuits, p. 266.
Fig. 27-3. Reprinted with permission from Popular Electronics, 6/93, p. 63. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 28

Fig. 28-1. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 65.
Fig. 28-2. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 65.

Chapter 29

Fig. 29-1. Reprinted with permission from 73 Amateur Radio Today, 3/92, p. 44.
Fig. 29-2. Reprinted with permission from 73 Amateur Radio Today, 1/92
Fig. 29-3. Reprinted with permission from 73 Amateur Radio Today, 5/90, p. 80.
Fig. 29-4. Reprinted with permission from 73 Amateur Radio Today, 9/90, p. 9.
Fig. 29-5. Reprinted with permission from Popular Electronics, 11/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 30

Fig. 30-1. Reprinted with permission from Popular Electronics, Fact Card, 226, (c) Copyright Gernsback Publications, Inc.
Fig. 30-2. Reprinted with permission from Popular Electronics, Fact Card, 226, (c) Copyright Gernsback Publications, Inc.
Fig. 30-3. Reprinted with permission from Popular Electronics, Fact Card, 226, (c) Copyright Gernsback Publications, Inc.
Fig. 30-4. Reprinted with permission from Popular Electronics, Fact Card, 227. (c) Copyright Gernsback Publications, Inc.
Fig. 30-5. Reprinted with permission from Popular Electronics, Fact Card, 227. (c) Copyright Gernsback Publications, Inc.
Fig. 30-6. Reprinted with permission from Popular Electronics, Fact Card, 227. (c) Copyright Gernsback Publications, Inc.
Fig. 30-7. Reprinted with permission from Popular Electronics, Fact Card, 227. (c) Copyright Gernsback Publications, Inc.
Fig. 30-8. Reprinted with permission from Popular Electronics, Fact Card, 228. (c) Copyright Gernsback Publications, Inc.
Fig. 30-9. Reprinted with permission from Popular Electronics, Fact Card, 228. (c) Copyright Gernsback Publications, Inc.
Fig. 30-10. Reprinted with permission from Popular Electronics, Fact Card, 228. (c) Copyright Gernsback Publications, Inc.
Fig. 30-11. Reprinted with permission from Popular Electronics, Fact Card, 228. (c) Copyright Gernsback Publications, Inc.

Fig. 30-12. Reprinted with permission from Popular Electronics, Fact Card, 225. (c) Copyright Gernsback Publications, Inc.

Fig. 30-13. Reprinted with permission from Popular Electronics, Fact Card, 224. (c) Copyright Gernsback Publications, Inc.

Fig. 30-14. Reprinted with permission from Popular Electronics, Fact Card, 223. (c) Copyright Gernsback Publications, Inc.

Fig. 30-15. Reprinted with permission from Popular Electronics, Fact Card, 231. (c) Copyright Gernsback Publications, Inc.

Fig. 30-16. Reprinted with permission from Popular Electronics, Fact Card, 231. (c) Copyright Gernsback Publications, Inc.

Fig. 30-17. Reprinted with permission from Electronic Design, 2/93, p. 75.

Fig. 30-18. Reprinted with permission from Electronics Now, 8/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-19. Reprinted with permission from Electronics Now, 8/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-20. Reprinted with permission from Electronics Now, 4/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-21. Reprinted with permission from Electronics Now, 8/93, p. 70. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-22. Reprinted with permission from Popular Electronics, 6/92, p. 68. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 30-23. Reprinted with permission from Electronics Now, 8/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-24. Reprinted with permission from Popular Electronics, 6/92, p. 67. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 30-25. Reprinted with permission from Electronics Now, 8/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-26. Reprinted with permission from Electronics Now, 8/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-27. Reprinted with permission from Linear Technology Corporation, 1993, Advertisement, Circle No. 51.

Fig. 30-28. Reprinted with permission from Electronics Now, 8/93, p. 70. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 30-29. William Sheets.

Fig. 30-30. William Sheets.

Fig. 30-31. William Sheets.

Fig. 30-32. William Sheets.

Fig. 30-33. William Sheets.

Fig. 30-34. William Sheets.

Fig. 30-35. Reprinted with permission from Electronic Design, 7/92, p. 62.

Fig. 30-36. Reprinted with permission from 303 Circuits, p. 185.

Fig. 30-37. Reprinted with permission from National Semiconductor, Linear Edge, Summer 1992.

Fig. 30-38. Reprinted with permission from Popular Electronics, Fact Card 224. (c) Copyright Gernsback Publications, Inc.

Fig. 30-39. Reprinted with permission from Popular Electronics, Fact Card 231. (c) Copyright Gernsback Publications, Inc.

Fig. 30-40. Reprinted with permission from Popular Electronics, Fact Card 242. (c) Copyright Gernsback Publications, Inc.

Fig. 30-41. William Sheets.

Chapter 31

Fig. 31-1. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 159.

Fig. 31-2. Reprinted with permission from Popular Electronics, 12/91, p. 80. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 31-3. William Sheets.

Fig. 31-4. William Sheets.

Fig. 31-5. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 10. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 31-6. William Sheets.

Fig. 31-7. Reprinted with permission from Radio-Electronics, 11/89, p. 12. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 31-8. Reprinted with permission from R-E Experimenters Handbook, p. 28.

Chapter 32

Fig. 32-1. Reprinted with permission from Electronic Engineering, 9/89, p. 30.

Chapter 33

Fig. 33-1. Reprinted with permission from Electronic Design, 7/93, p. 76.

Fig. 33-2. Reprinted with permission from Popular Electronics, Fact Card 254. (c) Copyright Gernsback Publications, Inc.

Fig. 33-3. Reprinted with permission from 73 Amateur Radio Today, 1/92, p. 28.

Fig. 33-4. Reprinted with permission from Popular Electronics, Fact Card 253. (c) Copyright Gernsback Publications, Inc.

Fig. 33-5. Reprinted with permission from Popular Electronics, 11/91, p. 22. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 33-6. Reprinted with permission from Elektor Electronics, 3/92, p. 58.

Fig. 33-7. William Sheets.

Fig. 33-8. William Sheets.

Fig. 33-9. Reprinted with permission from Popular Electronics, Fact Card 230. (c) Copyright Gernsback Publications, Inc.

Fig. 33-10. William Sheets.

Fig. 33-11. Reprinted with permission from Popular Electronics, Fact Card 257. (c) Copyright Gernsback Publications, Inc.

Fig. 33-12. Reprinted with permission from Popular Electronics, Fact Card 243. (c) Copyright Gernsback Publications, Inc.

Fig. 33-13. Reprinted with permission from Electronic Design, 7/93, p. 76.

Fig. 33-14. William Sheets.

Fig. 33-15. Reprinted with permission from Popular Electronics, Fact Card 258. (c) Copyright Gernsback Publications, Inc.

Chapter 34

Fig. 34-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 33.

Fig. 34-2. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 47. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 34-3. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 36-37. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 35

Fig. 35-1. Reprinted with permission from Popular Electronics, 11/93, p. 33. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 35-2. Reprinted with permission from Electronics Now, 7/93, p. 40. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 36

Fig. 36-1. Reprinted with permission from Popular Electronics, Fact Card, 268. (c) Copyright Gernsback Publications, Inc.

Chapter 37

Fig. 37-1. Reprinted with permission from Popular Electronics, 7/92, pp. 42-43. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 37-2. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 59. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 38

Fig. 38-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 81.

Fig. 38-2. Reprinted with permission from Popular Electronics, 7/93, p. 75. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 39

Fig. 39-1. Reprinted with permission from Popular Electronics, 6/92, p. 39. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 39-2. Reprinted with permission from Popular Electronics, 1/92, p. 24. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 39-3. Reprinted with permission from Radio-Electronics, 10/89, p. 43. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 39-4. Reprinted with permission from Popular Electronics, 1/92, p. 24. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 39-5. William Sheets.

Fig. 39-6. William Sheets.

Fig. 39-7. Reprinted with permission from Popular Electronics, 12/93, p. 32. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 39-8. Reprinted with permission from Popular Electronics, 12/93, p. 32. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 39-9. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 75-77. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 39-10. Reprinted with permission from Radio-Electronics, 10/89, p. 43. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 39-11. Reprinted with permission from Popular Electronics, 3/93, p. 43. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 39-12. Reprinted with permission from Electronics Now, 5/93, p. 12. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 39-13. Reprinted with permission from Electronics Now, 3/93, p. 83. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 40

Fig. 40-1. Reprinted with permission from Popular Electronics, 4/92, p. 88. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 40-2. Reprinted with permission from Popular Electronics, 4/92, pp. 70-71. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 41

Fig. 41-1. Reprinted with permission from National Semiconductor, Linear Edge, Spring 1992.

Fig. 41-2. Reprinted with permission from National Semiconductor, Linear Edge, Summer 1992.

Chapter 42

Fig. 42-1. Reprinted with permission from Popular Electronics, Fact Card 255. (c) Copyright Gernsback Publications, Inc.

Chapter 43

Fig. 43-1. Reprinted with permission from Popular Electronics, 4/92, p. 67. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 43-2. Reprinted with permission from Popular Electronics, 8/91, p. 75. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 43-3. Reprinted with permission from Popular Electronics, 8/92, p. 76. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 44

Fig. 44-1. Reprinted with permission from Radio-Electronics, 2/92, p. 66. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 44-2. Reprinted with permission from Popular Electronics, Fact Card 265. (c) Copyright Gernsback Publications, Inc.

Fig. 44-3. Reprinted with permission from 73 Amateur Radio Today, 4/89, p. 87. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 44-4. Reprinted with permission from Popular Electronics, 6/93, p. 55. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 44-5. Reprinted with permission from Radio-Electronics, 7/90, p. 64. (c) Copyright Gernsback Publications, Inc., 1990.

Chapter 45

Fig. 45-1. Reprinted with permission from Electronics Now, 10/92, p. 76. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 45-2. William Sheets.

Fig. 45-3. Reprinted with permission from Popular Electronics, Fact Card 267. (c) Copyright Gernsback Publications, Inc.

Fig. 45-4. Reprinted with permission from Popular Electronics, Fact Card 266. (c) Copyright Gernsback Publications, Inc.

Chapter 46

Fig. 46-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 14.

Chapter 47

Fig. 47-1. Reprinted with permission from Electronic Design, 10/93, p. 73.

Fig. 47-2. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 62.

Fig. 47-3. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 122.

Fig. 47-4. William Sheets.

Fig. 47-5. Reprinted with permission from PE Hobbyist handbook, 1991, pp. 42-43. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 47-6. Reprinted with permission from Popular Electronics, 6/93, p. 78. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 48

Fig. 48-1. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 89-90. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 49

Fig. 49-1. Reprinted with permission from Popular Electronics, Application Circuit 215. (c) Copyright Gernsback Publications, Inc.

Fig. 49-2. Reprinted with permission from Popular Electronics, Application Circuit 215. (c) Copyright Gernsback Publications, Inc.

Fig. 49-3. Reprinted with permission from Electronics Now, 12/92, p. 60. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 49-4. Reprinted with permission from Electronics Now, 12/92, p. 59. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 49-5. William Sheets.

Fig. 49-6. William Sheets.

Fig. 49-7. Reprinted with permission from Popular Electronics, 3/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 49-8. Reprinted with permission from Popular

Electronics, 3/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 50

Fig. 50-1. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 79-80. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 50-2. Reprinted with permission from PE Hobbyist Handbook, 1990, pp. 45-47. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 50-3. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 64.

Fig. 50-4. Reprinted with permission from 303 Circuits, p. 238.

Fig. 50-5. Reprinted with permission from Electronics Design, 5/92, p. 93.

Fig. 50-6. Reprinted with permission from Radio-Electronics, 7/90, p. 65. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 50-7. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 82. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 50-8. Reprinted with permission from Radio-Electronics, 7/90, p. 64. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 50-9. Reprinted with permission from 303 Circuits, p. 251.

Fig. 50-10. Reprinted with permission from Popular Electronics, 11/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 50-11. Reprinted with permission from Elektor Electronics, 12/91, p. 87.

Fig. 50-12. Reprinted with permission from Radio-Electronics, 10/89, p. 12. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 50-13. Reprinted with permission from Electronic Design, 5/92, p. 93.

Fig. 50-14. Reprinted with permission from 303 Circuits, p. 63.

Chapter 51

Fig. 51-1. Reprinted with permission from Popular Electronics, 8/92, p. 73. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 51-2. William Sheets.

Fig. 51-3. William Sheets.

Fig. 51-4. William Sheets.

Fig. 51-5. William Sheets.

Fig. 51-6. William Sheets.

Fig. 51-7. William Sheets.

Fig. 51-8. William Sheets.

Fig. 51-9. Reprinted with permission from Popular Electronics, 2/92, p. 90. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 51-10. William Sheets.

Fig. 51-11. William Sheets.

Fig. 51-12. William Sheets.

Fig. 51-13. Reprinted with permission from Popular Electronics, 3/93, p. 43. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 51-14. William Sheets.

Fig. 51-15. William Sheets.

Chapter 52

Fig. 52-1. Reprinted with permission from PE Hobbyist Handbook, pp. 93-94.

Fig. 52-2. Reprinted with permission from Popular Electronics, 6/92, p. 33. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 53

Fig. 53-1. Reprinted with permission from Popular Electronics, 12/91, p. 22. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 53-2. Reprinted with permission from Popular Electronics, 12/91, p. 18. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 54

Fig. 54-1. Reprinted with permission from Electronic Design, 2/93, p. 83.

Fig. 54-2. Reprinted with permission from 73 Amateur Radio Today, 11/92, p. 34.

Fig. 54-3. Reprinted with permission from Popular Electronics, Fact Card 258. (c) Copyright Gernsback Publications, Inc.

Chapter 55

Fig. 55-1. Reprinted with permission from Radio-Electronics, 12/91, pp. 31-36. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 55-2. Reprinted with permission from Radio-Electronics, 12/91, p. 48. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 55-3. William Sheets.

Fig. 55-4. Reprinted with permission from Popular Electronics, 2/92, pp. 53-54. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-5. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 39.

Fig. 55-6. Reprinted with permission from Radio-Electronics, 5/92, p. 52. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-7. Reprinted with permission from Electronics Now, 7/93, p. 45. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-8. Reprinted with permission from Radio-Electronics, 8/86, p. 42. (c) Copyright Gernsback Publications, Inc., 1986.

Fig. 55-9. Reprinted with permission from Popular Electronics, 4/92, p. 53. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-10. Reprinted with permission from Popular Electronics, 12/91, p. 26. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 55-11. Reprinted with permission from Popular Electronics, 9/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-12. William Sheets.

Fig. 55-13. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 101.

Fig. 55-14. Reprinted with permission from Popular Electronics, 3/93, p. 42. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-15. William Sheets.

Fig. 55-16. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 90.

Fig. 55-17. Reprinted with permission from Electronic Design, 4/93, p. 94.

Fig. 55-18. Reprinted with permission from Radio-Electronics, 12/91, p. 51. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 55-19. Reprinted with permission from Popular Electronics, 5/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-20. Reprinted with permission from 303 Circuits, p. 308.

Fig. 55-21. Reprinted with permission from Electronics Now, 12/92, p. 64. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-22. William Sheets.

Fig. 55-23. Reprinted with permission from Popular Electronics, 3/93, p. 42. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-24. Reprinted with permission from Popular Electronics, Fact Card 110. (c) Copyright Gernsback Publications, Inc.

Fig. 55-25. Reprinted with permission from Electronic Design, 4/93, p. 56.

Fig. 55-26. Reprinted with permission from Popular Electronics, Fact Card 221. (c) Copyright Gernsback Publications, Inc.

Fig. 55-27. Reprinted with permission from Popular Electronics, Fact Card 221. (c) Copyright Gernsback Publications, Inc.

Fig. 55-28. Reprinted with permission from Popular Electronics, 11/93, p. 42. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-29. Reprinted with permission from Popular Electronics, 9/93, p. 45. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-30. Reprinted with permission from 303 Circuits, p. 187.

Fig. 55-31. Reprinted with permission from 73 Amateur Radio Today, 5/92, p. 62.

Fig. 55-32. Reprinted with permission from Electronics Now, 8/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-33. Reprinted with permission from Popular Electronics, 11/91, p. 18. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 55-34. Reprinted with permission from R-E Experimenters Handbook, 1989, pp. 156-157.

Fig. 55-35. Reprinted with permission from Popular Electronics, 3/93, p. 75. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-36. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 31.

Fig. 55-37. Reprinted with permission from Electronic Design, 5/92, p. 92.

Fig. 55-38. Reprinted with permission from Popular Electronics, 9/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 55-39. Reprinted with permission from 73 Amateur Radio Today, 1/92, p. 38.

Fig. 55-40. Reprinted with permission from Popular Electronics, 3/93, p. 42. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-41. Reprinted with permission from 303 Circuits, p. 248.

Fig. 55-42. William Sheets.

Fig. 55-43. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 14. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 55-44. William Sheets.

Fig. 55-45. Reprinted with permission from Electronic Design, 3/93.

Fig. 55-46. Reprinted with permission from Popular Electronics, 9/93, p. 46. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 55-47. Reprinted with permission from Popular Electronics, Fact Card, 198. (c) Copyright Gernsback Publications, Inc.

Fig. 55-48. Reprinted with permission from Popular Electronics, Fact Card, 221. (c) Copyright Gernsback Publications, Inc.

Fig. 55-49. Reprinted with permission from Popular Electronics, 3/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 56

Fig. 56-1. Reprinted with permission from 303 Circuits, pp. 249-250.

Fig. 56-2. Reprinted with permission from Popular Electronics, 3/89, p. 69. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 56-3. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 71-72. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 57

Fig. 57-1. Reprinted with permission from Popular Electronics, 12/92, pp. 53-54. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 57-2. Reprinted with permission from Electronics Hobbyist Handbook, 1992, p. 93.

Fig. 57-3. Reprinted with permission from Popular Electronics, 3/93, p. 36. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 57-4. Reprinted with permission from Popular Electronics, 10/92, pp. 39-40. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 57-5. William Sheets.

Fig. 57-6. Reprinted with permission from 303 Circuits, p 265.

Fig. 57-7. Reprinted with permission from Popular Electronics, 11/93, p. 55. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 57-8. Reprinted with permission from R-E Experimenters Handbook, 1989, pp. 38-39.

Fig. 57-9. Reprinted with permission from Radio-Electronics, 1/92, p. 82. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 57-10. Reprinted with permission from Electronic Engineering, 11/89, pp. 21-22.

Fig. 57-11. Reprinted with permission from R-E Experimenters Handbook, pp. 118-120.

Fig. 57-12. Reprinted with permission from Electronic Design, 8/92, p. 70.

Fig. 57-13. Reprinted with permission from Electronics Now, 7/92, p. 10. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 57-14. Reprinted with permission from Popular Electronics, 12/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 57-15. Reprinted with permission from 73 Amateur Radio Today, 8/92, p. 48.

Fig. 57-16. Reprinted with permission from Maxim Engineering Journal, Volume 4, pp. 11-12.

Fig. 57-17. Reprinted with permission from Radio-Electronics, 10/89, p. 13. (c) Copyright Gernsback Publications, Inc., 1989.

Fig. 57-18. Reprinted with permission from Elektor Electronics, 3/92, p.20.

Fig. 57-19. Reprinted with permission from Popular Electronics, Fact Card 223. (c) Copyright Gernsback Publications, Inc.

Fig. 57-20. Reprinted with permission from Popular Electronics, Fact Card 257. (c) Copyright Gernsback Publications, Inc.

Fig. 57-21. Reprinted with permission from Popular Electronics, Fact Card 259. (c) Copyright Gernsback Publications, Inc.

Fig. 57-22. Reprinted with permission from Popular Electronics, 11/93, p. 80. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 57-23. Reprinted with permission from 73 Amateur Radio Today, 3/92, p. 8.

Fig. 57-24. Reprinted with permission from Electronic Design, 4/93, p. 93.

Fig. 57-25. Reprinted with permission from Popular Electronics, 11/91, p. 18. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 57-26. Reprinted with permission from Popular Electronics, 11/91, p. 18. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 57-27. Reprinted with permission from Electronic Design, 7/92, p. 59.

Fig. 57-28. Reprinted with permission from Popular Electronics, 6/93, p. 55. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 57-29. Reprinted with permission from Electronic Design, 10/93, p. 74.

Fig. 57-30. Reprinted with permission from 73 Amateur Radio Today, 4/89, p. 87.

Fig. 57-31. William Sheets.

Fig. 57-32. Reprinted with permission from Popular Electronics, 9/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 57-33. Reprinted with permission from Precision Monolithics Inc., 1981. Full Line Catalog, pp. 6-59.

Fig. 57-34. William Sheets.

Fig. 57-35. Reprinted with permission from Linear Databook, 1986, pp. 8-12.
Fig. 57-36. Reprinted with permission from QST, 3/89, p. 36.
Fig. 57-37. Reprinted with permission from 73 Amateur Radio Today, 2/93, p. 46. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 57-38. Reprinted with permission from 73 Amateur Radio Today, 2/93, p. 48. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 57-39. Reprinted with permission from Electronics Now, 10/92, p. 80. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 57-40. Reprinted with permission from Electronic Design, 5/92, p. 94.
Fig. 57-41. William Sheets.
Fig. 57-42. Reprinted with permission from QST, 3/89, p.35.
Fig. 57-43. Reprinted with permission from 73 Amateur Radio Today, 2/93, p. 46.
Fig. 57-44. Reprinted with permission from Popular Electronics, 3/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 58

Fig. 58-1. Reprinted with permission from Silicon Chip, p. 56.
Fig. 58-2. Reprinted with permission from Electronics Now, 10/93, p. 12. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 58-3. William Sheets.
Fig. 58-4. William Sheets.
Fig. 58-5. Reprinted with permission from Popular Electronics, Fact Card 264. (c) Copyright Gernsback Publications, Inc.

Chapter 59

Fig. 59-1. Reprinted with permission from RF Design, 3/93, pp. 92-93.
Fig. 59-2. William Sheets.
Fig. 59-3. Reprinted with permission from RF Design, 3/93, p. 92.

Chapter 60

Fig. 60-1. Reprinted with permission from Silicon Chip, p. 46.
Fig. 60-2. Reprinted with permission from Popular Electronics, 10/92, pp. 31-32. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 60-3. Reprinted with permission from Popular Electronics, 10/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 60-4. Reprinted with permission from R-E Experimenters Handbook, p.41.

Chapter 61

Fig. 61-1. Reprinted with permission from Popular Electronics, Fact Card 198. (c) Copyright Gernsback Publications, Inc.
Fig. 61-2. Reprinted with permission from Popular Electronics, 9/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 61-3. Reprinted with permission from PE Hobbyist Handbook, p. 12. (c) Copyright Gernsback Publications, Inc.

Chapter 62

Fig. 62-1. Reprinted with permission from Popular Electronics, 10/93, p. 31. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 63

Fig. 63-1. Reprinted with permission from Radio-Electronics, 7/90, p. 66. (c) Copyright Gernsback Publications, Inc., 1990.
Fig. 63-2. Reprinted with permission from Radio-Electronics, 2/92, p. 12. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 63-3. Reprinted with permission from Popular Electronics, 6/93, p. 73.
Fig. 63-4. Reprinted with permission from Apex Microtechnology Corporation.
Fig. 63-5. Reprinted with permission from Popular Electronics, 3/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 63-6. Reprinted with permission from Popular Electronics, 6/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 64

Fig. 64-1. William Sheets.

Chapter 65

Fig. 65-1. William Sheets.
Fig. 65-2. William Sheets.
Fig. 65-3. William Sheets.
Fig. 65-4. William Sheets.
Fig. 65-5. William Sheets.
Fig. 65-6. William Sheets.
Fig. 65-7. William Sheets.
Fig. 65-8. Reprinted with permission from Popular Electronics, Fact Card 259. (c) Copyright Gerns-

back Publications, Inc.

Fig. 65-9. Reprinted with permission from Electronics Now, 10/92, p. 69. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 65-10. Reprinted with permission from Popular Electronics, Fact Card 268. (c) Copyright Gernsback Publications, Inc.

Chapter 66

Fig. 66-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 17.

Fig. 66-2. Reprinted with permission from Electronics Now, 6/93, p. 47. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 66-3. Reprinted with permission from Elektor Electronics, 3/92, p. 15.

Fig. 66-4. Reprinted with permission from Electronics Now, 11/92, p. 63. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 66-5. Reprinted with permission from Elektor Electronics, 3/92, p. 14.

Fig. 66-6. Reprinted with permission from Radio-Electronics, 7/90, p. 66. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 66-7. Reprinted with permission from Radio-Electronics, 3/92, p. 77. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 67

Fig. 67-1. Reprinted with permission from 303 Circuits, p. 312.

Chapter 68

Fig. 68-1. Reprinted with permission from Popular Electronics, Fact Card 242. (c) Copyright Gernsback Publications, Inc.

Fig. 68-2. Reprinted with permission from 73 Amateur Radio Today, 11/92, p. 12.

Fig. 68-3. Reprinted with permission from 73 Amateur Radio Today, 11/92, p. 12.

Fig. 68-4. Reprinted with permission from 73 Amateur Radio Today, 11/92, p. 12.

Chapter 69

Fig. 69-1. Reprinted with permission from Electronic Design, 1/93, p. 116.

Fig. 69-2. Reprinted with permission from Electronic Design, 4/93, p. 93.

Fig. 69-3. Reprinted with permission from Popular Electronics, Fact Card 253. (c) Copyright Gernsback Publications, Inc.

Fig. 69-4. Reprinted with permission from Electronic Design, 3/89, p. 100.

Fig. 69-5. William Sheets.

Fig. 69-6. Reprinted with permission from Electronic Design, 1/93, p. 63.

Fig. 69-7. Reprinted with permission from Maxim Engineering Journal, Volume 3, p. 17.

Fig. 69-8. Reprinted with permission from Maxim Engineering Journal, Volume 3, p. 28.

Chapter 70

Fig. 70-1. Reprinted with permission from Popular Electronics, 3/93, p. 45. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 70-2. Reprinted with permission from Electronic Engineering, 8/93, p. 18.

Fig. 70-3. Reprinted with permission from Popular Electronics, 10/92, pp. 55-56. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 70-4. Reprinted with permission from Popular Electronics, 10/92, p. 56. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 70-5. William Sheets.

Fig. 70-6. Reprinted with permission from Popular Electronics, 9/93, p. 76. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 70-7. Reprinted with permission from Electronics Now, 11/92, p. 14. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 70-8. William Sheets.

Chapter 71

Fig. 71-1. Reprinted with permission from 73 Amateur Radio Today, 2/93, p. 48.

Fig. 71-2. Reprinted with permission from Popular Electronics, Fact Card 229. (c) Copyright Gernsback Publications, Inc.

Fig. 71-3. Reprinted with permission from Popular Electronics, Fact Card 230. (c) Copyright Gernsback Publications, Inc.

Fig. 71-4. Reprinted with permission from QST, 2/89, pp. 33-35.

Fig. 71-5. Reprinted with permission from Popular Electronics, 11/91, p. 21. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 71-6. Reprinted with permission from Radio-Electronics, 12/91, p. 12. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 71-7. William Sheets.

Fig. 71-8. William Sheets.

Fig. 71-9. William Sheets.

Fig. 71-10. William Sheets.

Fig. 71-11. Reprinted with permission from 303 Circuits, p. 323.
Fig. 71-12. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 59.
Fig. 71-13. Reprinted with permission from Popular Electronics, 2/92, pp. 29-31. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 71-14. Reprinted with permission from Popular Electronics, 12/93, p. 70. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 71-15. Reprinted with permission from Popular Electronics, 12/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 71-16. William Sheets.
Fig. 71-17. Reprinted with permission from Popular Electronics, Fact Card 260. (c) Copyright Gernsback Publications, Inc.
Fig. 71-18. William Sheets.
Fig. 71-19. Reprinted with permission from Popular Electronics, 8/93, p. 79. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 71-20. Reprinted with permission from Popular Electronics, 12/93, p. 68. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 71-21. Reprinted with permission from Popular Electronics, 11/91, p. 21. (c) Copyright Gernsback Publications, Inc., 1991.
Fig. 71-22. Reprinted with permission from Popular Electronics, 12/93, p. 70. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 72

Fig. 72-1. Reprinted with permission from Radio-Electronics, 6/92, p. 60. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 72-2. Reprinted with permission from Popular Electronics, 12/91, p. 77. (c) Copyright Gernsback Publications, Inc., 1991.
Fig. 72-3. William Sheets.
Fig. 72-4. William Sheets.
Fig. 72-5. Reprinted with permission from Electronics Now, 7/92, p. 88. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 73

Fig. 73-1. Reprinted with permission from PE Hobbyist Handbook, 1992, p. 61. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 73-2. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 69-70. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 74

Fig. 74-1. William Sheets.
Fig. 74-2. William Sheets.
Fig. 74-3. Reprinted with permission from Electronic Design, 1/93, p. 62. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 75

Fig. 75-1. Reprinted with permission from PE Hobbyist Handbook, 1992, p. 41. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 75-2. Reprinted with permission from Popular Electronics, Fact Card 198. (c) Copyright Gernsback Publications, Inc.
Fig. 75-3. Reprinted with permission from Popular Electronics, 4/92, p. 31. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 75-4. Reprinted with permission from Electronics Now, 11/92, p. 32. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 75-5. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 38. (c) Copyright Gernsback Publications, Inc., 1991.
Fig. 75-6. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 73. (c) Copyright Gernsback Publications, Inc., 1991.
Fig. 75-7. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 54. (c) Copyright Gernsback Publications, Inc., 1991.
Fig. 75-8. Reprinted with permission from Popular Electronics, Fact Card 198. (c) Copyright Gernsback Publications, Inc.

Chapter 76

Fig. 76-1. Reprinted with permission from Popular Electronics, 5/92, p. 60. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 76-2. Reprinted with permission from Popular Electronics, 5/92, p. 60. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 76-3. Reprinted with permission from Apex Microtechnology Corporation.
Fig. 76-4. Reprinted with permission from Apex Microtechnology Corporation.
Fig. 76-5. Reprinted with permission from Popular Electronics, 5/92, p. 60. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 77

Fig. 77-1. Reprinted with permission from PE Hobbyist Handbook, 1990, p. 92. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 77-2. Reprinted with permission from Popular Electronics, 6/93, p. 77. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 77-3. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 62.

Fig. 77-4. Reprinted with permission from Popular Electronics, 6/93, p. 77. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 77-5. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 62.

Fig. 77-6. Reprinted with permission from Elektor Electronics, 12/91, p. 94.

Fig. 77-7. Reprinted with permission from Popular Electronics, 7/93, p. 76. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 77-8. Reprinted with permission from Electronic Design, 6/93, p. 76.

Chapter 78

Fig. 78-1. Reprinted with permission from Electronics Now, 10/93, p. 53. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-2. Reprinted with permission from R-E Experimenters Handbook, p. 60.

Fig. 78-3. Reprinted with permission from Electronic Design, 2/93, p. 71.

Fig. 78-4. Reprinted with permission from Maxim Engineering Journal, Volume 3, p. 16.

Fig. 78-5. Reprinted with permission from Linear Technology, Design Note 72.

Fig. 78-6. Reprinted with permission from Popular Electronics, 6/93, p. 48. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-7. Reprinted with permission from Popular Electronics, 3/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 78-8. Reprinted with permission from Popular Electronics, 6/93, p. 77. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-9. Reprinted with permission from 303 Circuits, p. 283.

Fig. 78-10. Reprinted with permission from Electronics Now, 12/92, p. 66. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 78-11. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 60.

Fig. 78-12. Reprinted with permission from Popular Electronics, 5/92, p. 73. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 78-13. Reprinted with permission from PE Hobbyist Handbook, 1993, p. 93. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-14. William Sheets.

Fig. 78-15. Reprinted with permission from PE Hobbyist Handbook, 1991. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 78-16. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 28-29. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 78-17. Reprinted with permission from Popular Electronics, 11/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-18. Reprinted with permission from Silicon Chip, pp. 63-64.

Fig. 78-19. Reprinted with permission from Maxim Engineering Journal, Volume 4, p. 19.

Fig. 78-20. Reprinted with permission from National Semiconductor, Linear Edge, Issue #5.

Fig. 78-21. Reprinted with permission from National Semiconductor, Linear Edge, Issue #5.

Fig. 78-22. Reprinted with permission from 73 Amateur Radio Today, 3/92, p. 54.

Fig. 78-23. Reprinted with permission from Popular Electronics, 1/92, p. 37. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 78-24. Reprinted with permission from Popular Electronics, 11/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-25. Reprinted with permission from Silicon Chip, p. 10.

Fig. 78-26. Reprinted with permission from Electronic Design, 2/93, p. 72.

Fig. 78-27. Reprinted with permission from Electronic Design, 8/93, p. 84.

Fig. 78-28. Reprinted with permission from Linear Technology, Design Note 68.

Fig. 78-29. Reprinted with permission from Popular Electronics, Fact Card 260. (c) Copyright Gernsback Publications, Inc.

Fig. 78-30. Reprinted with permission from Popular Electronics, 11/93, p. 54.

Fig. 78-31. Reprinted with permission from Popular Electronics, 8/93, p. 88. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-32. Reprinted with permission from Popular Electronics, 3/93, p. 74. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-33. Reprinted with permission from PE Hobbyist Handbook, 1993, p. 61. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-34. Reprinted with permission from Electronic Design, 2/93, pp. 75-76.

Fig. 78-35. Reprinted with permission from Electronics Now, 10/93, p. 54. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 78-36. Reprinted with permission from Electronic Design, 4/93, p. 54.

Fig. 78-37. Reprinted with permission from National Semiconductor, Linear Edge, Spring 1992.

Chapter 79

Fig. 79-1. Reprinted with permission from Popular Electronics, 9/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 79-2. Reprinted with permission from Popular Electronics, 11/91, p. 22. (c) Copyright Gernsback Publications, Inc., 1991.

Chapter 80

Fig. 80-1. Reprinted with permission from Popular Electronics, 3/92, p. 42. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 80-2. Reprinted with permission from Radio-Electronics, 12/91, p. 63. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 80-3. Reprinted with permission from Popular Electronics, 9/93, p. 69. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 80-4. Reprinted with permission from Elektor Electronics, 12/91, p. 73.

Fig. 80-5. Reprinted with permission from Elektor Electronics, 12/91, p. 72.

Fig. 80-6. Reprinted with permission from Elektor Electronics, 12/91, p. 85.

Fig. 80-7. Reprinted with permission from Electronics Now, 12/92, p. 45. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 80-8. Reprinted with permission from Electronic Design, 7/93, p. 87.

Fig. 80-9. Reprinted with permission from Popular Electronics, 9/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 80-10. Reprinted with permission from Electronics Now, 12/92, p. 46. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 80-11. Reprinted with permission from Popular Electronics, 10/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 80-12. Reprinted with permission from Silicon Chip, p. 64.

Fig. 80-13. Reprinted with permission from Popular Electronics, 9/93, p. 69. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 81

Fig. 81-1. Reprinted with permission from PE Hobbyist Handbook, 1991, pp. 85-86. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 81-2. Reprinted with permission from Popular Electronics, Fact Card 270. (c) Copyright Gernsback Publications, Inc.

Chapter 82

Fig. 82-1. William Sheets.

Fig. 82-2. William Sheets.

Fig. 82-3. Reprinted with permission from PE Hobbyists Handbook, 1990, p. 120. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 82-4. Reprinted with permission from Electronic Design 1/93, p. 61.

Fig. 82-5. William Sheets.

Fig. 82-6. William Sheets.

Fig. 82-7. William Sheets.

Chapter 83

Fig. 83-1. Reprinted with permission from 73 Amateur Radio Today, 10/91, p. 8.

Fig. 83-2. Reprinted with permission from Electronics Now, 10/92, p. 37. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 83-3. Reprinted with permission from Popular Electronics, 10/93, p. 74. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 83-4. Reprinted with permission from Popular Electronics, 8/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 83-5. Reprinted with permission from Popular Electronics, 8/93, p. 70. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 83-6. Reprinted with permission from Integrated Circuits Data Book, 3/85, pp. 5-16.

Fig. 83-7. Reprinted with permission from Popular Electronics, 3/93, p. 79. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 83-8. Reprinted with permission from 73 Amateur Radio Today, 8/93, p. 32.

Fig. 83-9. Reprinted with permission from Popular

Electronics, 6/92, p. 55. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 83-10. Reprinted with permission from Popular Electronics, 8/93, p. 32. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 83-11. Reprinted with permission from QST, 2/89, p. 34.

Fig. 83-12. Reprinted with permission from Popular Electronics, 6/92, p. 57. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 84

Fig. 84-1. Reprinted with permission from Electronic Design, 8/92, p. 69.

Fig. 84-2. Reprinted with permission from Radio-Electronics, 5/92, p. 47. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 84-3. William Sheets.

Fig. 84-4. Reprinted with permission from Electronics Now, 3/93, p. 69. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 84-5. William Sheets.

Chapter 85

Fig. 85-1. Reprinted with permission from Electronics Now, 11/92, p. 53. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 85-2. Reprinted with permission from Electronics Now, 11/92, p. 54. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 85-3. Reprinted with permission from Popular Electronics, 8/93, p. 56. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 85-4. Reprinted with permission from R-E Experimenters Handbook, 1991, p. 30.

Fig. 85-5. Reprinted with permission from Popular Electronics, 3/93, p. 45. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 85-6. Reprinted with permission from Popular Electronics, 8/93, p. 53. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 85-7. Reprinted with permission from Popular Electronics, 3/93, p. 45. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 86

Fig. 86-1. Reprinted with permission from Popular Electronics, 6/93, p. 55. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-2. Reprinted with permission from Popular Electronics, 6/93, p. 56. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-3. Reprinted with permission from Popular Electronics, Fact Card 225. (c) Copyright Gernsback Publications, Inc.

Fig. 86-4. William Sheets.

Fig. 86-5. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 156.

Fig. 86-6. Reprinted with permission from Practical Wireless, 6/91, p. 34.

Fig. 86-7. Reprinted with permission from R-E Experimenters Handbook, p. 33.

Fig. 86-8. Reprinted with permission from Electronic Design, 6/93, p. 83.

Fig. 86-9. Reprinted with permission from 73 Amateur Radio Today, 5/90, p. 78.

Fig. 86-10. Reprinted with permission from 73 Amateur Radio Today, 5/90, p. 78.

Fig. 86-11. Reprinted with permission from Popular Electronics, Fact Card 241. (c) Copyright Gernsback Publications, Inc.

Fig. 86-12. Reprinted with permission from 73 Amateur Radio Today, 11/91, pp. 52-56. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 86-13. Reprinted with permission from 73 Amateur Radio Today, 10/92, p. 20.

Fig. 86-14. Reprinted with permission from Popular Electronics, 11/93, p. 81. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-15. Reprinted with permission from 73 Amateur Radio, 5/90, p. 78.

Fig. 86-16. Reprinted with permission from Popular Electronics, 8/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-17. Reprinted with permission from Popular Electronics, Fact Card 262. (c) Copyright Gernsback Publications, Inc.

Fig. 86-18. Reprinted with permission from 73 Amateur Radio, 5/90, p. 77.

Fig. 86-19. Reprinted with permission from Popular Electronics, 8/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-20. Reprinted with permission from Popular Electronics, 3/93, p. 47. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-21. Reprinted with permission from Popular Electronics, 9/93, p. 83. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 86-22. Reprinted with permission from 73 Amateur Radio Today, 2/93, p. 60.

Fig. 86-23. William Sheets.
Fig. 86-24. Reprinted with permission from 73 Amateur Radio Today, 5/90, p. 31.
Fig. 86-25. Reprinted with permission from Popular Electronics, 6/93, p. 54. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 86-26. William Sheets.

Chapter 87

Fig. 87-1. Reprinted with permission from QST, 6/91, p. 18.
Fig. 87-2. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 30.
Fig. 87-3. Reprinted with permission from Popular Electronics, 6/92, p. 56. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 87-4. Reprinted with permission from 73 Amateur Radio Today, 3/92, p. 16.
Fig. 87-5. Reprinted with permission from Popular Electronics, 7/93, p. 80. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 88

Fig. 88-1. Reprinted with permission from Popular Electronics, 11/93, p. 73. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 88-2. Reprinted with permission from Maxim Journal, Vol. 3., p. 22.

Chapter 89

Fig. 89-1. Reprinted with permission from PE Hobbyist Handbook, 1990, p. 21. (c) Copyright Gernsback Publications, Inc., 1990.

Chapter 90

Fig. 90-1. Reprinted with permission from Popular Electronics, 5/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 90-2. Reprinted with permission from Popular Electronics, 5/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 91

Fig. 91-1. Reprinted with permission from Electronic Design, 11/92, p. 62.
Fig. 91-2. Reprinted with permission from Popular Electronics, 12/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 91-3. Reprinted with permission from Popular Electronics, Fact Card 256. (c) Copyright Gernsback Publications, Inc.
Fig. 91-4. Reprinted with permission from Electronics Now, 6/93, p. 14. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 91-5. Reprinted with permission from Popular Electronics, Fact Card 256. (c) Copyright Gernsback Publications, Inc.
Fig. 91-6. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 160.
Fig. 91-7. Reprinted with permission from Electronic Design, 10/93, p. 74.
Fig. 91-8. Reprinted with permission from Popular Electronics, 11/91, p. 22. (c) Copyright Gernsback Publications, Inc., 1991.
Fig. 91-9. Reprinted with permission from Maxim Engineering Journal, Vol. 4, p. 15.

Chapter 92

Fig. 92-1. Reprinted with permission from R-E Experimenters Handbook, 1992, p. 98.
Fig. 92-2. Reprinted with permission from Electronic Design, 6/93, p. 82.
Fig. 92-3. Reprinted with permission from Popular Electronics, 10/92, p. 58. (c) Copyright Gernsback Publications, Inc., 1992.
Fig. 92-4. Reprinted with permission from Popular Electronics, Fact Card 201. (c) Copyright Gernsback Publications, Inc.
Fig. 92-5. Reprinted with permission from 73 Amateur Radio Today, 11/91, p. 11.
Fig. 92-6. Reprinted with permission from Popular Electronics, 6/93, p. 59. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 92-7. Reprinted with permission from Radio-Electronics, 1/92, p. 12.
Fig. 92-8. Reprinted with permission from Popular Electronics, Fact Card 255. (c) Copyright Gernsback Publications, Inc.
Fig. 92-9. Reprinted with permission from Popular Electronics, 6/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 92-10. Reprinted with permission from Electronics Now, 12/93, p. 39. (c) Copyright Gernsback Publications, Inc., 1993.
Fig. 92-11. Reprinted with permission from 303 Circuits.

Chapter 93

Fig. 93-1. Reprinted with permission from Popular Electronics, 1/92, p. 43. (c) Copyright Gernsback

Publications, Inc., 1992.

Fig. 93-2. Reprinted with permission from Electronic Design, 8/93, p. 81.

Fig. 93-3. William Sheets.

Fig. 93-4. William Sheets.

Fig. 93-5. Reprinted with permission from Popular Electronics, 12/91, p. 81. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 93-6. William Sheets.

Fig. 93-7. William Sheets.

Fig. 93-8. Reprinted with permission from PE Hobbyist, 1991, p. 77. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 93-9. William Sheets.

Fig. 93-10. William Sheets.

Fig. 93-11. Reprinted with permission from Popular Electronics, 2/92, p. 67. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 93-12. Reprinted with permission from Popular Electronics, 2/92, p. 66. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 93-13. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 155.

Fig. 93-14. Reprinted with permission from National Semiconductor, Linear Applications Handbook.

Fig. 93-15. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 161.

Fig. 93-16. Reprinted with permission from 303 Circuits, p. 257 (#221).

Chapter 94

Fig. 94-1. Reprinted with permission from Popular Electronics, Fact Card 223. (c) Copyright Gernsback Publications, Inc.

Fig. 94-2. William Sheets.

Fig. 94-3. Reprinted with permission from Popular Electronics, Fact Card 221. (c) Copyright Gernsback Publications, Inc.

Fig. 94-4. Reprinted with permission from Popular Electronics, Fact Card 243. (c) Copyright Gernsback Publications, Inc.

Fig. 94-5. William Sheets.

Fig. 94-6. William Sheets.

Fig. 94-7. Reprinted with permission from Popular Electronics, Fact Card 263. (c) Copyright Gernsback Publications, Inc.

Chapter 95

Fig. 95-1. Reprinted with permission from Popular Electronics, 3/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 95-2. Reprinted with permission from Popular Electronics, 3/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 95-3. Reprinted with permission from Popular Electronics, 3/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 96

Fig. 96-1. Reprinted with permission from Radio-Electronics, 6/92, p. 71. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 96-2. Reprinted with permission from R-E Experimenters Handbook, 1989, pp. 12-15.

Fig. 96-3. Reprinted with permission from Electronics Now, 7/92, p. 33. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 96-4. Reprinted with permission from Popular Electronics, Fact Card 261. (c) Copyright Gernsback Publications, Inc.

Fig. 96-5. Reprinted with permission from 303 Circuits, p. 49.

Fig. 96-6. Reprinted with permission from Electronics Now, 3/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 96-7. Reprinted with permission from 303 Circuits, p. 41.

Fig. 96-8. Reprinted with permission from Popular Electronics, Fact Card 261. (c) Copyright Gernsback Publications, Inc.

Fig. 96-9. Reprinted with permission from Popular Electronics, Fact Card 241. (c) Copyright Gernsback Publications, Inc.

Fig. 96-10. Reprinted with permission from Popular Electronics, Fact Card 262. (c) Copyright Gernsback Publications, Inc.

Chapter 97

Fig. 97-1. Reprinted with permission from Radio-Electronics, 12/90, pp. 72-73.

Fig. 97-2. Reprinted with permission from Popular Electronics, 6/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 97-3. Reprinted with permission from National Semiconductor, Linear Edge, Summer 1992.

Fig. 97-4. Reprinted with permission from Popular Electronics, 12/92, p. 32. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 97-5. Reprinted with permission from Popular Electronics, 9/93, p. 70. (c) Copyright Gernsback

Publications, Inc., 1993.

Fig. 97-6. Reprinted with permission from Popular Electronics, 6/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 97-7. Reprinted with permission from Popular Electronics, 6/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 97-8. Reprinted with permission from Maxim Engineering Journal, Vol. 4. p. 10.

Fig. 97-9. Reprinted with permission from Electronics Now, 6/93, p. 14. (c) Copyright Gernsback Publications, Inc.

Fig. 97-10. Reprinted with permission from Popular Electronics, 6/91, p. 71. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 97-11. Reprinted with permission from Popular Electronics, 6/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 97-12. Reprinted with permission from Popular Electronics, 9/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 97-13. Reprinted with permission from Popular Electronics, 6/93, p. 72. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 98

Fig. 98-1. Reprinted with permission from Radio-Electronics, 6/90, p. 71.

Fig. 98-2. Reprinted with permission from Radio-Electronics, 6/90, p. 71.

Chapter 99

Fig. 99-1. Reprinted with permission from Popular Electronics, 7/92, pp. 60-61. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 99-2. William Sheets.

Chapter 100

Fig. 100-1. Reprinted with permission from Radio-Electronics, 7/90, p. 8.

Fig. 100-2. Reprinted with permission from Radio-Electronics, 10/89, p. 8.

Fig. 100-3. Reprinted with permission from Radio-Electronics, 11/91, p. 59.

Fig. 100-4. Reprinted with permission from Popular Electronics, 4/92, p. 38. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-5. Reprinted with permission from Popular Electronics, p. 75. (c) Copyright Gernsback Publications, Inc.

Fig. 100-6. Reprinted with permission from Popular Electronics, 9/92, pp. 38-40. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-7. Reprinted with permission from 303 Circuits, 226, p. 263.

Fig. 100-8. Reprinted with permission from Popular Electronics, 8/92, p. 76. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-9. Reprinted with permission from Popular Electronics, 9/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-10. Reprinted with permission from Popular Electronics, 12/91, p. 53. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 100-11. Reprinted with permission from Radio-Electronics, 1/93, p. 43. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 100-12. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 26.

Fig. 100-13. Reprinted with permission from Popular Electronics, 9/93, p. 33. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 100-14. Reprinted with permission from Popular Electronics, 11/93, p. 38. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 100-15. Reprinted with permission from Electronic Design, 8/93, p. 86.

Fig. 100-16. Reprinted with permission from Electronics Now, 5/93, p. 47. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 100-17. Reprinted with permission from Silicon Chip, p. 62.

Fig. 100-18. Reprinted with permission from Popular Electronics, 12/93, p. 62. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 100-19. Reprinted with permission from Radio-Electronics, 3/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-20. Reprinted with permission from Radio-Electronics, 2/92, p. 81. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-21. Reprinted with permission from Popular Electronics, 2/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-22. Reprinted with permission from Electronics Now, 11/92, p. 45. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-23. Reprinted with permission from Popular Electronics, 7/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 100-24. Reprinted with permission from Popular Electronics, 11/92, p. 72. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 101

Fig. 101-1. Reprinted with permission from Electronic Design, 5/92, p. 94.

Fig. 101-2. Reprinted with permission from 303 Circuits, #229, pp. 264-265.

Fig. 101-3. Reprinted with permission from Electronic Design, 11/92, p. 62. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 101-4. Reprinted with permission from Intersil, Component Data Catalog, 1987, pp. 6-10.

Fig. 101-5. Reprinted with permission from Radio-Electronics, 3/90, p. 50. (c) Copyright Gernsback Publications, Inc., 1990.

Fig. 101-6. Reprinted with permission from 303 Circuits, #228, p. 364.

Fig. 101-6. Reprinted with permission from Radio-Electronics, 3/92, p. 50. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 101-7. Reprinted with permission from Radio-Electronics, 6/92, p. 54. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 101-8. Reprinted with permission from Radio-Electronics.

Chapter 102

Fig. 102-1. Reprinted with permission from Electronics Now, 10/92, p. 43. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 102-2. Reprinted with permission from Popular Electronics, 5/92, p. 75. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 102-3. William Sheets.

Fig. 102-4. William Sheets.

Fig. 102-5. William Sheets.

Fig. 102-6. William Sheets.

Fig. 102-7. William Sheets.

Fig. 102-8. Reprinted with permission from PE Hobbyist Handbook, 1991, p. 57. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 102-9. Reprinted with permission from Popular Electronics, 12/92, p. 68. (c) Copyright Gernsback Publications, Inc., 1992.

Chapter 103

Fig. 103-1. Reprinted with permission from Practical Wireless, 2/91, p. 49.

Fig. 103-2. Reprinted with permission from 73 Amateur Radio, 7/88, p. 14.

Chapter 104

Fig. 104-1. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 101.

Fig. 104-2. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 101.

Fig. 104-3. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 100.

Chapter 105

Fig. 105-1. Reprinted with permission from Popular Electronics, 8/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 105-2. Reprinted with permission from Popular Electronics, Fact Card 266. (c) Copyright Gernsback Publications, Inc.

Fig. 105-3. Reprinted with permission from Popular Electronics, 8/92, p. 74. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 105-4. Reprinted with permission from Popular Electronics, 12/93, p. 71. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 105-5. Reprinted with permission from Popular Electronics, 7/92, p. 70. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 105-6. Reprinted with permission from Hands-On Electronics, 9/87, p. 88.

Chapter 106

Fig. 106-1. Reprinted with permission from Electronics Now, 10/92, p. 36. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 106-2. Reprinted with permission from 73 Amateur Radio Today, 10/91, pp. 14-22.

Fig. 106-3. William Sheets.

Fig. 106-4. Reprinted with permission from Popular Electronics, 8/92, p. 46. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 106-5. Reprinted with permission from Radio-Electronics, 11/91, p. 85. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 106-6. Reprinted with permission from 73 Amateur Radio Today, 11/92, p. 8.

Fig. 106-7. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 24.

Fig. 106-8. Reprinted with permission from Electronics Hobbyist Handbook, 1993, p. 52.

Fig. 106-9. Reprinted with permission from QST, 10/89, p. 25.

Fig. 106-10. Reprinted with permission from 73 Amateur Radio Today, 4/92, p. 36.

Fig. 106-11. Reprinted with permission from 73 Amateur Radio Today, 7/92, p. 20.

Fig. 106-12. Reprinted with permission from 73 Amateur Radio Today, 4/92, p. 25.

Fig. 106-13. Reprinted with permission from R-E Experimenters Handbook, 1989, p. 158.

Fig. 106-14. Reprinted with permission from 73 Amateur Radio Today, 4/93, p. 53.

Chapter 107

Fig. 107-1. Reprinted with permission from Radio-Electronics, 11/91, pp. 49-57. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 107-2. Reprinted with permission from Radio-Electronics, 11/91, p. 49. (c) Copyright Gernsback Publications, Inc., 1991.

Fig. 107-3. Reprinted with permission from Electronics Now, 3/93, p. 33. (c) Copyright Gernsback Publications, Inc., 1993.

Chapter 108

Fig. 108-1. Reprinted with permission from Radio-Electronics, 6/92, p. 61. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 108-2. Reprinted with permission from Radio-Electronics, 6/92, p. 59. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 108-3. Reprinted with permission from 303 Circuits, p. 307.

Fig. 108-4. Reprinted with permission from Spec-Com, 5/91, p. 15.

Fig. 108-5. Reprinted with permission from Linear Technology Design Note #57.

Fig. 108-6. Reprinted with permission from Linear Technology Design Note #57.

Fig. 108-7. Reprinted with permission from Linear Technology Design Note #57.

Fig. 108-8. Reprinted with permission from Linear Technology Design Note #57.

Fig. 108-9. Reprinted with permission from Linear Technology Application Note #57.

Fig. 108-10. Reprinted with permission from Linear Technology Application Note #57.

Fig. 108-11. Reprinted with permission from Radio-Electronics, 4/92, p. 64. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 108-12. Reprinted with permission from 303 Circuits, #303, p. 332.

Fig. 108-13. Reprinted with permission from Radio-Electronics, 6/92, p. 59. (c) Copyright Gernsback Publications, Inc., 1992.

Fig. 108-14. Reprinted with permission from Electronics Now, 8/93, p. 39. (c) Copyright Gernsback Publications, Inc., 1993.

Fig. 108-15. Reprinted with permission from 303 Circuits, #300, p. 331.

Fig. 108-16. Reprinted with permission from Popular Electronics, Fact Card 268. (c) Copyright Gernsback Publications, Inc.

Chapter 109

Fig. 109-1. Reprinted with permission from Electronic Engineering, 9/89, p. 28.

Fig. 109-2. William Sheets.

Fig. 109-3. Reprinted with permission from Popular Electronics, Fact Card 269. (c) Copyright Gernsback Publications, Inc.

Fig. 109-4. Reprinted with permission from Popular Electronics, Fact Card #241. (c) Copyright Gernsback Publications, Inc.

Fig. 109-5. Reprinted with permission from Popular Electronics, Fact Card 224. (c) Copyright Gernsback Publications, Inc.

Fig. 109-6. William Sheets.

Chapter 110

Fig. 110-1. Reprinted with permission from 303 Circuits, p. 280.

Fig. 110-2. William Sheets.

Chapter 111

Fig. 111-1. William Sheets.

Fig. 111-2. William Sheets.

Fig. 111-3. William Sheets.

Chapter 112

Fig. 112-1. William Sheets.

Fig. 112-2. Reprinted with permission from Popular Electronics, Fact Card 259. (c) Copyright Gernsback Publications, Inc.

Index

Numbers preceded by a "I,", "II," "III,", "IV," or "V" are from *Encyclopedia of Electronic Circuits* Vol. I, II, III, IV, or V respectively.

A

B

D

E

F

G

H

I

J

K

L

N

O

Q

R

S

T

U

V

WITH CUMULATIVE INDEX

Rudolf F. Graf
&
William Sheets

Encyclopedia of ELECTRONIC CIRCUITS

Volume 6

TAB
The Electronics Authority

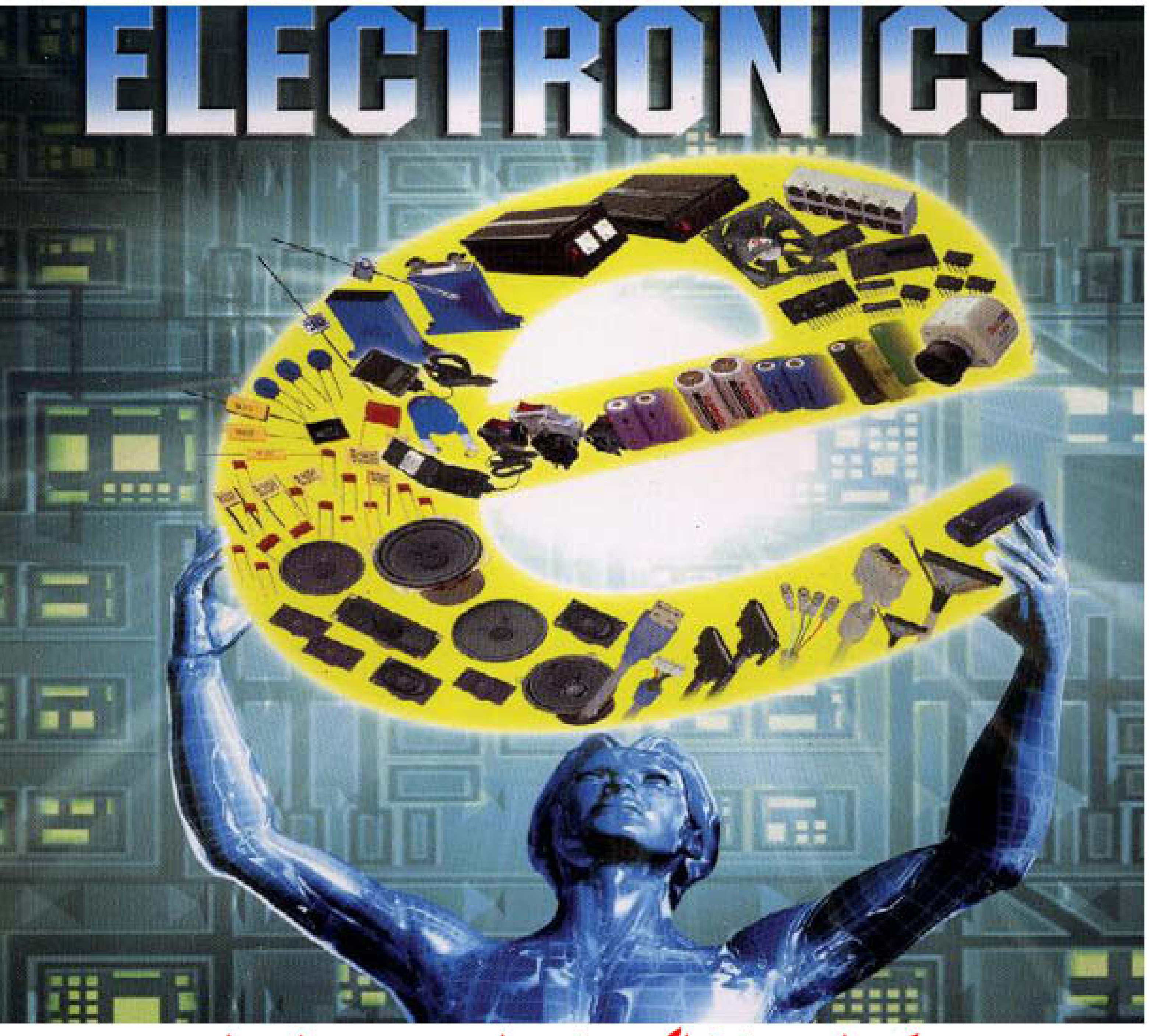
Encyclopedia of

ELECTRONIC CIRCUITS

Volume 6

To Russell
May you always be as self-confident as you are today.

Patent notice


ELECTRONICS

Encyclopedia of

ELECTRONIC CIRCUITS

Volume 6

Rudolf F. Graf
&
William Sheets

McGraw-Hill

New York San Francisco Washington, D.C. Auckland Bogotá
Caracas Lisbon London Madrid Mexico City Milan
Montreal New Delhi San Juan Singapore
Sydney Tokyo Toronto

McGraw-Hill

A Division of The **McGraw-Hill** *Companies*

Published by The McGraw-Hill Companies, Inc.

pbk 1 2 3 4 5 6 7 8 9 0 FGR/FGR 9 0 0 9 8 7 6

hc 1 2 3 4 5 6 7 8 9 0 FGR/FGR 9 0 0 9 8 7 6

Library of Congress Cataloging-in-Publication Data
(Revised for vol. 6)

Graf, Rudolf F.
The encyclopedia of electronics circuits

Authors for v. 6– : Ruldolf F. Graf & William Sheets.
Includes bibliographical references and indexes.
1. Electronic circuits—Encyclopedias. I. Sheets, William. II. Title
TK7867G66 1985 621.3815 84-26772
ISBN 0-8306-0938-5 (v. 1)
ISBN 0-8306-1938-0 (pbk. : v. 1)
ISBN 0-8306-3138-0 (pbk. : v 2)
ISBN 0-8306-3138-0 (v. 2)
ISBN 0-8306-3348-0 (pbk. : v 3)
ISBN 0-8306-7348-2 (v. 3)
ISBN 0-8306-3895-4 (pbk. : v 4)
ISBN 0-8306-3896-2 (v. 4)
ISBN 0-07-011077-8 (pbk. : v 5)
ISBN 0-07-011076-X (v. 5)
ISBN 0-07-011275-4 (v. 6)
ISBN 0-07-011276-2 (pbk. : v 6)

McGraw-Hill books are available at special quantity discounts to use as premiums and sales promotions, or for use in corporate training programs. For more information, please write to the Director of Special Sales, McGraw-Hill, 11 West 19th Street, New York, NY 10011. Or contact your local bookstore.

Acquisitions editor: Roland S. Phelps
Editorial team: Lori Flaherty, Executive Editor
Andrew Yoder, Book Editor
Joann Woy, Indexer
Production team: Katherine G. Brown, Director
Rose McFarland, Desktop Operator
Nancy Mickley, Proofreading
Design team: Jaclyn J. Boone, Designer
Katherine Lukaszewicz, Associate Designer

EL1
0112762

Contents

Introduction

The enthusiastic reception of the first five volumes of *The Encyclopedia of Electronic Circuits* prompted the authors to produce this volume—the sixth in the popular series.

Taken together, the six volumes contain approximately 6000 circuits—by far the largest and broadest collection of practical electronic circuits available anywhere.

As in the other volumes, the 1000+ circuits presented here are arranged alphabetically, by category. All circuits in this volume, as well as those from the previous five volumes, are included in the index, which now has approximately 6000 entries.

We express sincere appreciation to the many electronic industry sources and publishers who graciously allowed us to utilize some of their materials. Their cooperation is gratefully acknowledged.

Once again, it gives us great pleasure to extend our sincerest thanks to Loretta Gonsalves-Battiste, a fine lady whose skill at the computer and willingness to work long and hard made on-time delivery of the manuscript for this book possible.

Rudolf F. Graf & William Sheets
September 1995

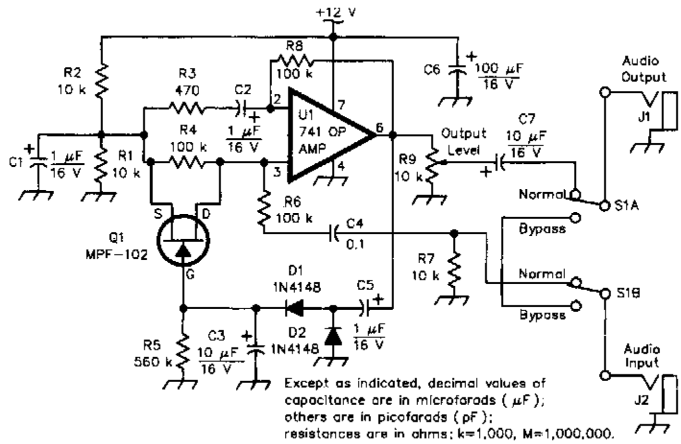
1

AGC and ALC Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

AGC Audio Preamp
3-MHz Low-Noise AGC System
IF AGC Network
Audio Leveler

AGC AUDIO PREAMP

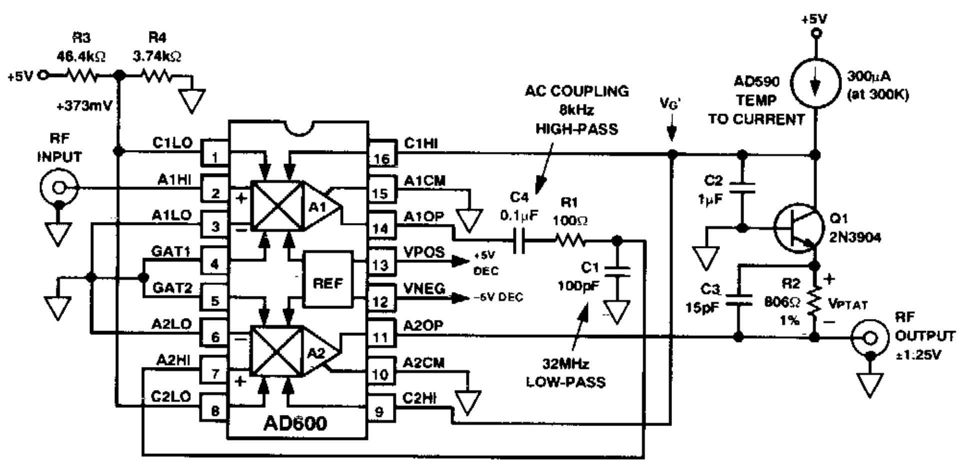


QST *Fig. 1-1*

The circuit uses an easily obtained 741 op amp set for an internal gain of about 200. A portion of the op amp's output signal is rectified by the 1N4148 diodes, then filtered and fed to the gate of the FET input shunting circuit. As the output rises, more and more input shunting takes place. That is, more of the input signal is bypassed, effectively keeping the output level constant.

The circuit offers a 100:1 limiting action. The input level can change over a 100:1 ratio with little or no effect on the output level. The output level itself can be set from less than unity all the way up to nearly the gain of the amplifier, making the circuit usable in other applications as well.

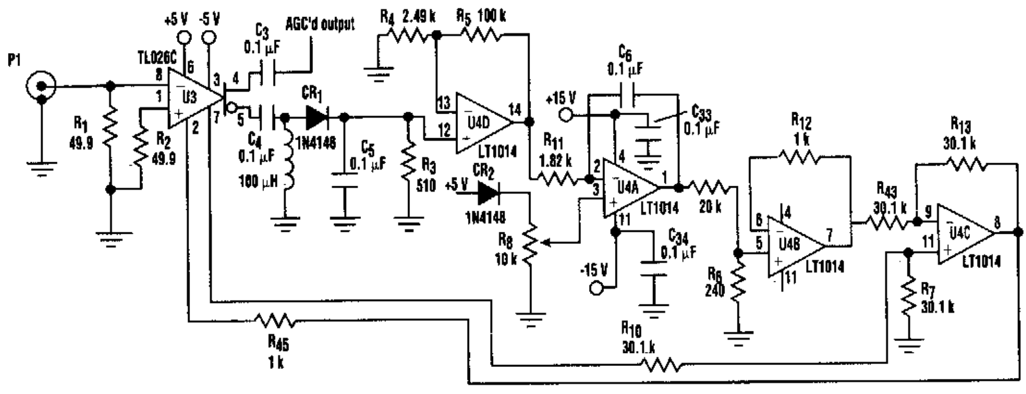
3-MHz LOW-NOISE AGC SYSTEM



ANALOG DIALOG *Fig. 1-2*

The AD600 dual voltage-controlled amplifier in this circuit provides a 3-MHz AGC system with 80-dB range.

IF AGC NETWORK

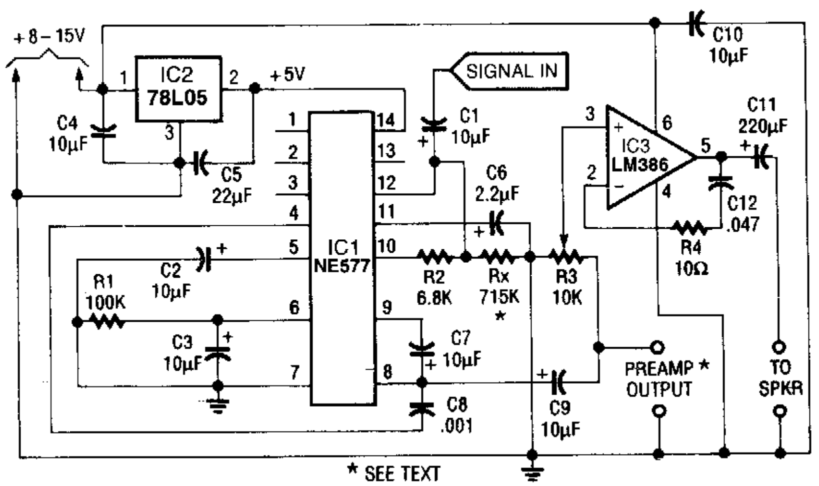


ELECTRONIC DESIGN

Fig. 1-3

A simple IF AGC circuit that features wide dynamic range and excellent linearity can be achieved with two chips: TI's TL026C voltage-controlled amplifier IC and Linear Technology's LT1014 (or any other similar basic quad op amp).

AUDIO LEVELER



1994 EXPERIMENTERS HANDBOOK

Fig. 1-4

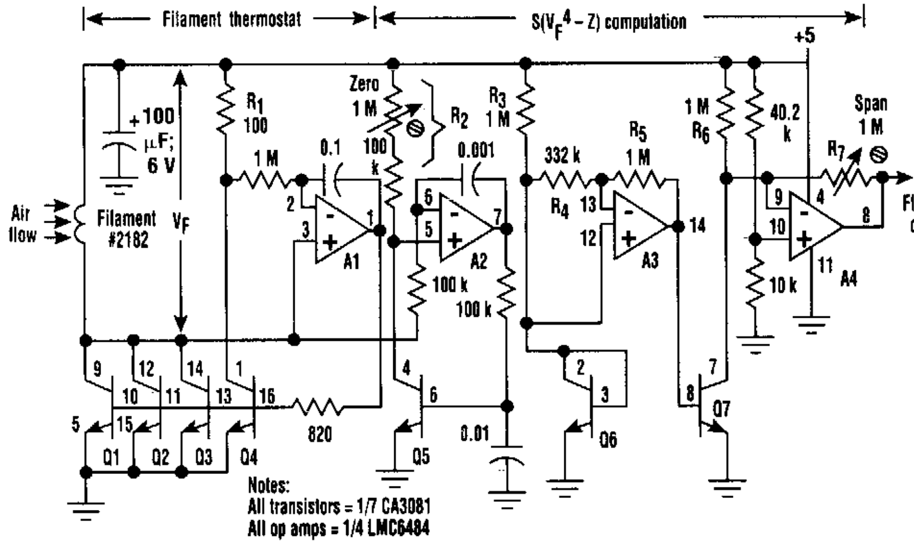
A low power programmable compandor chip, the Signetics NE577 IC is used. Incoming audio is compressed, rectified and conditioned so that the input signal level always remains about the noise level. The compressor is an ALC circuit that outputs a constant level and the expander part of the IC is not used.

2

Air-Flow Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

HOT-WIRE ANEMOMETER

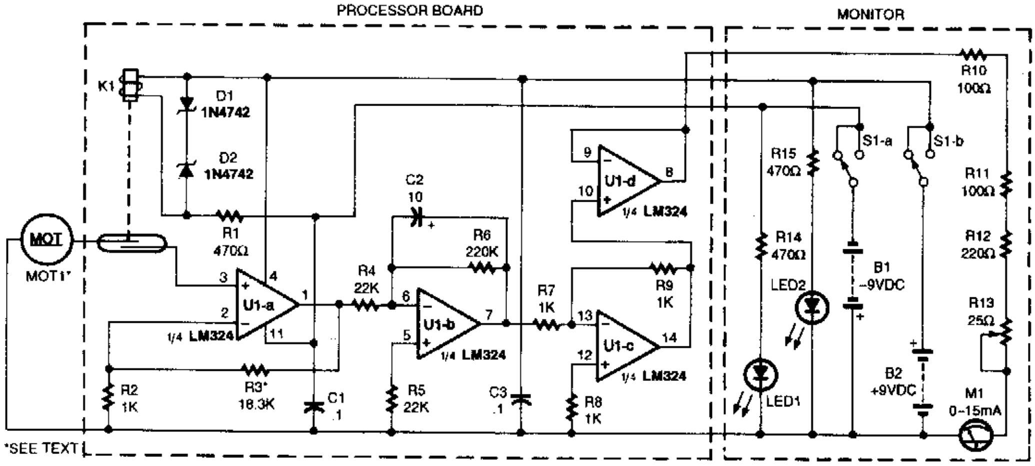


ELECTRONIC DESIGN *Fig. 2-1*

An anemometer can be realized by utilizing the inherent transconductance match of transistors in the array, instead of passive series resistances, to control filament current. As a result, as A1 serves the collector current of Q4 and thereby the voltage across R1, it simultaneously adjusts the filament (a 2182-type incandescent lamp denuded of its glass envelope) voltage, V_f. The ratio of the filament to R1 current is stably maintained by the identical temperature and operating points of Q1 through Q4. The net result is that A1 drives the filament temperature to the value that causes filament resistance to equal $R1/3 = 33\ \Omega$. This is about double the cold resistance of the filament and therefore, assuming tungsten wire with a 0.0045/degree coefficient of resistance, represents a filament operating temperature of around 230°C. This is hot enough that moderate changes in ambient temperature are unimportant factors in filament power demand, but not so hot as to cause the filament to burn.

Rail-to-rail input amplifier A2 continuously serves the collector current of Q5 to V_f/R_2, making the V_{be} of LQ5 a logarithmic function of V_f. A3 multiplies this log by 4 and applies the product of Q7. Q7 does the antilog function so that its collector current is proportional to the fourth power of V_f. Thus, by King's law, it's proportional to air speed in the vicinity of the filament. This current is offset and scaled by A4 to produce a voltage output that, thanks to the rail-to-rail output capability of the LMC6484, can range from 0.01 to 4.99 V. Full-scale air speed can be adjusted, using R7, to any value in the range of 1 to 10 meters/s.

ELECTRONIC ANEMOMETER

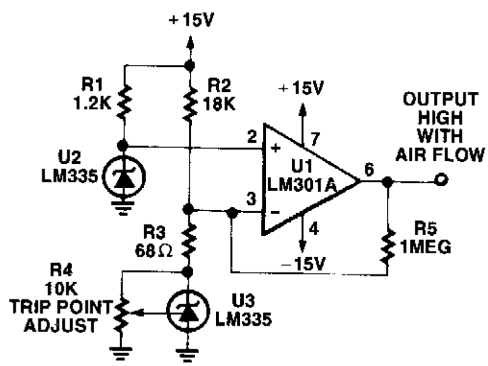


POPULAR ELECTRONICS

Fig. 2-2

A motor used as a generator is used as a transducer to generate a dc voltage that is proportional to wind speed. K1 prevents the transducer voltages from being applied to the circuit if no dc power is present. U1A through U1D is a dc amplifier, integrator, and buffer. This circuit drives the meter M1. The processor board is mounted in a housing along with the generator M1.

AIR FLOW DETECTOR



POPULAR ELECTRONICS

Fig. 2-3

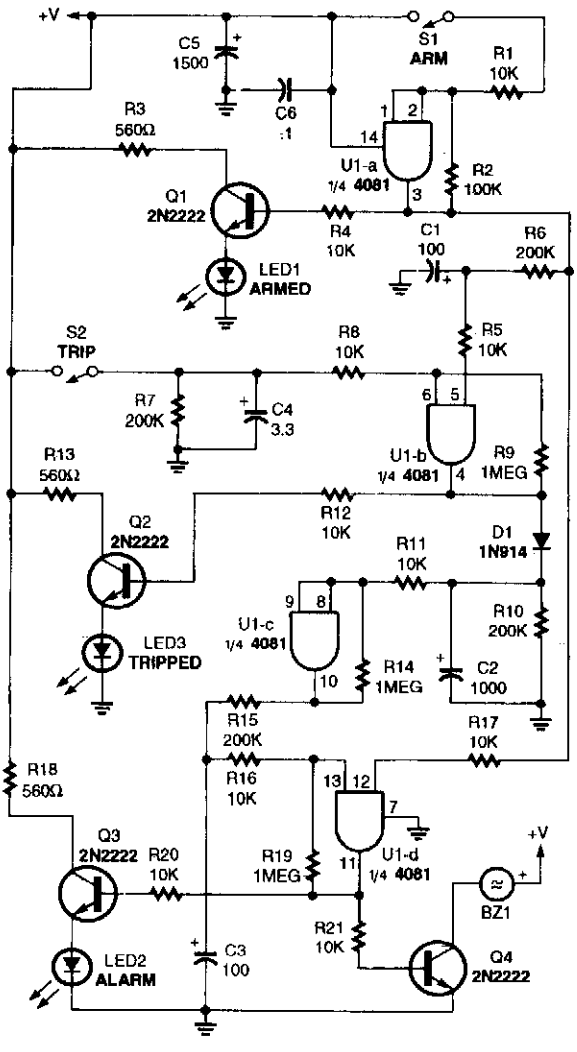
The self heating of a semiconductor that is cooled by airflow is used as a sensing method.

3

Alarm and Security Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

BURGLAR ALARM CIRCUIT



POPULAR ELECTRONICS *Fig. 3-1*

This alarm circuit is built around a single 4081 (CMOS) quad AND gate. It offers an exit and entry delay (around automatically reset two minutes after tripping, provided that the trip input is not left high).

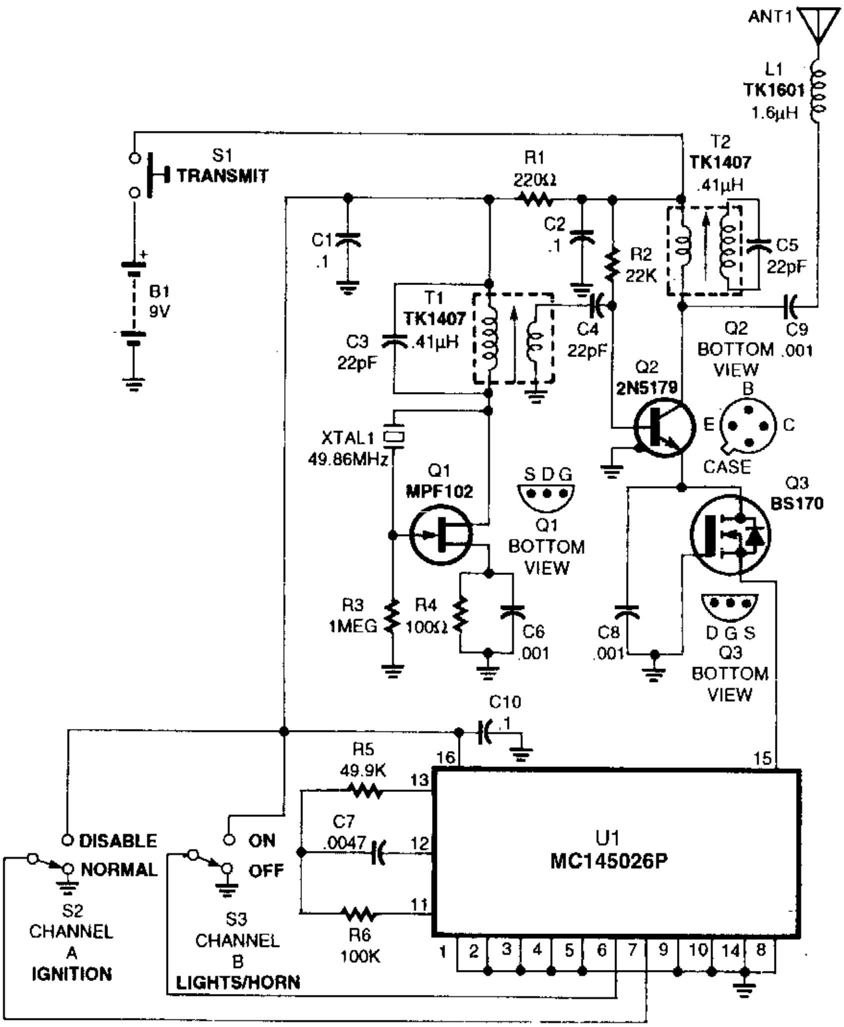
The arming switch must go high to arm or low to disarm. After arming, U1-a begins to charge C1 via R6. Around 20 seconds later (after the exit delay), C1 has a sufficient charge to produce a high at the pin-5 input of U1-b. Also, when the circuit is armed, Q1 is turned on to indicate arming, and one input of U1-d is brought high.

After the exit delay times out, if the trip input opens, it causes an output on gate U1-b. Transistor Q1 is turned on, lighting the trip indicator (LED3), C2 instantly charges, and the output of U1-c goes high. At that point, C3 begins charging to provide the entry delay.

After 20 seconds, C3 has sufficient charge to produce a high at pin 13 of U1-d. That forces U1-d's output high, tuning Q3 and Q4 on, which activates the alarm indicator (LED2) and sounder (BZ1), respectively. If disarmed after a trip pulse, but before the 20-second, entry delay time out, pin 12 of U1-d goes low, so the gate's output does not go high and the alarm does not sound.

Components C2 and R10 hold U1-c on for around 2 minutes and 20 seconds to provide the two-minute alarm. After C2's charge drops below half of the supply voltage, U1-c's output goes low, awaiting another trip pulse to set it off again.

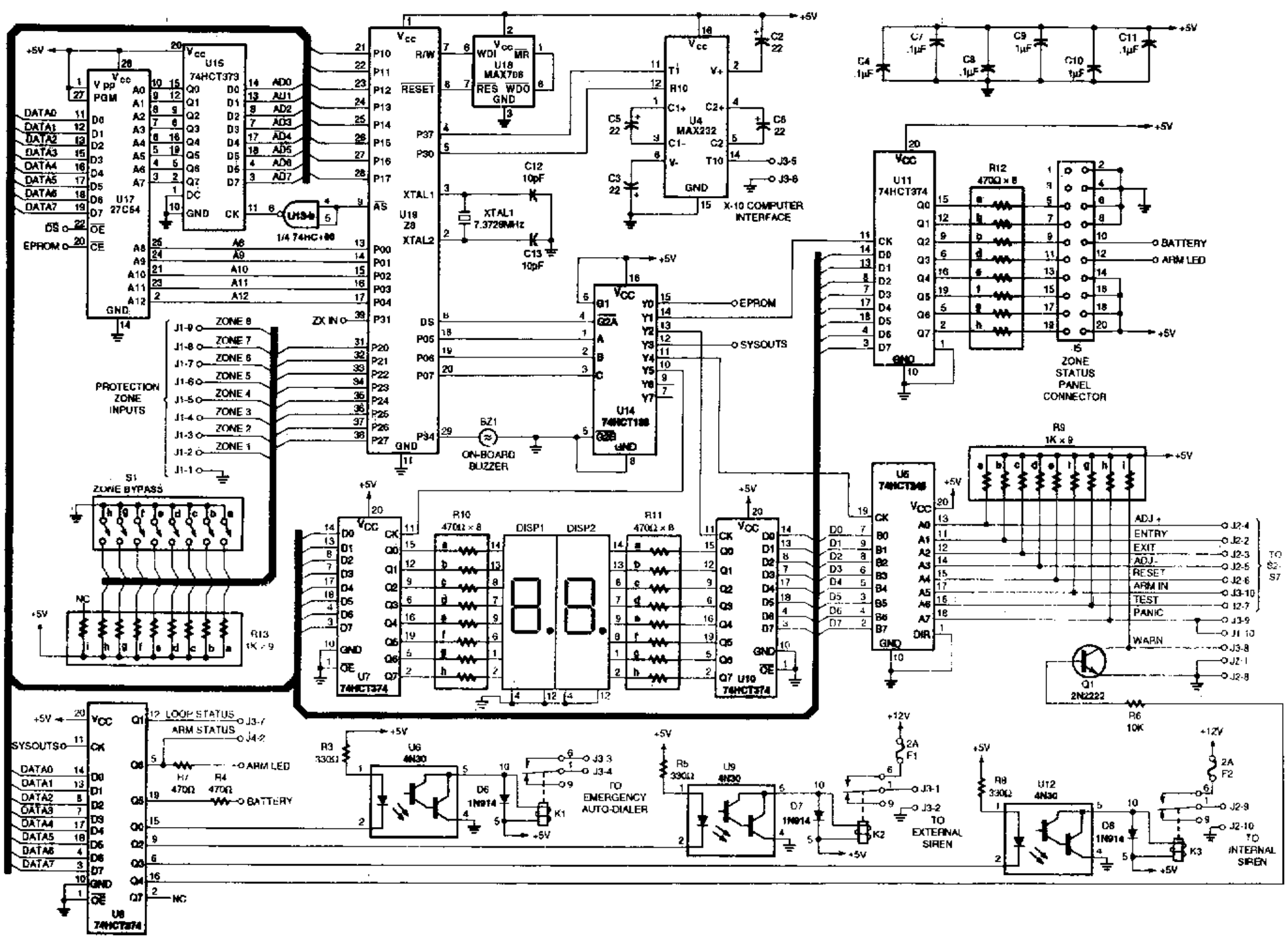
AUTO SECURITY SYSTEM TRANSMITTER

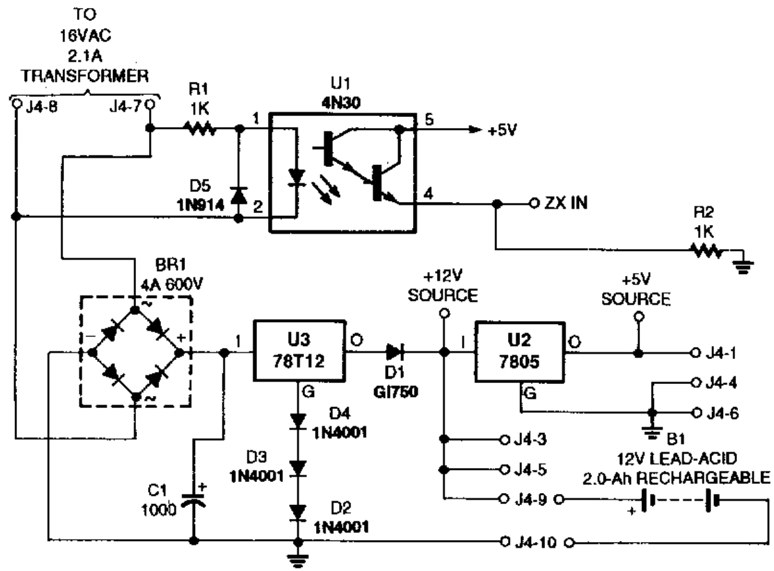


POPULAR ELECTRONICS *Fig. 3-2*

This transmitter operates at 49 MHz and uses an M145026 programmable digital encoder to generate a unique digital code, depending on the positions of S2 and S3, to control ignition and lights or horn. Q1 is the oscillator, Q2 the power amplifier. The antenna is a 36-inch whip or wire antenna.

HOME SECURITY SYSTEM





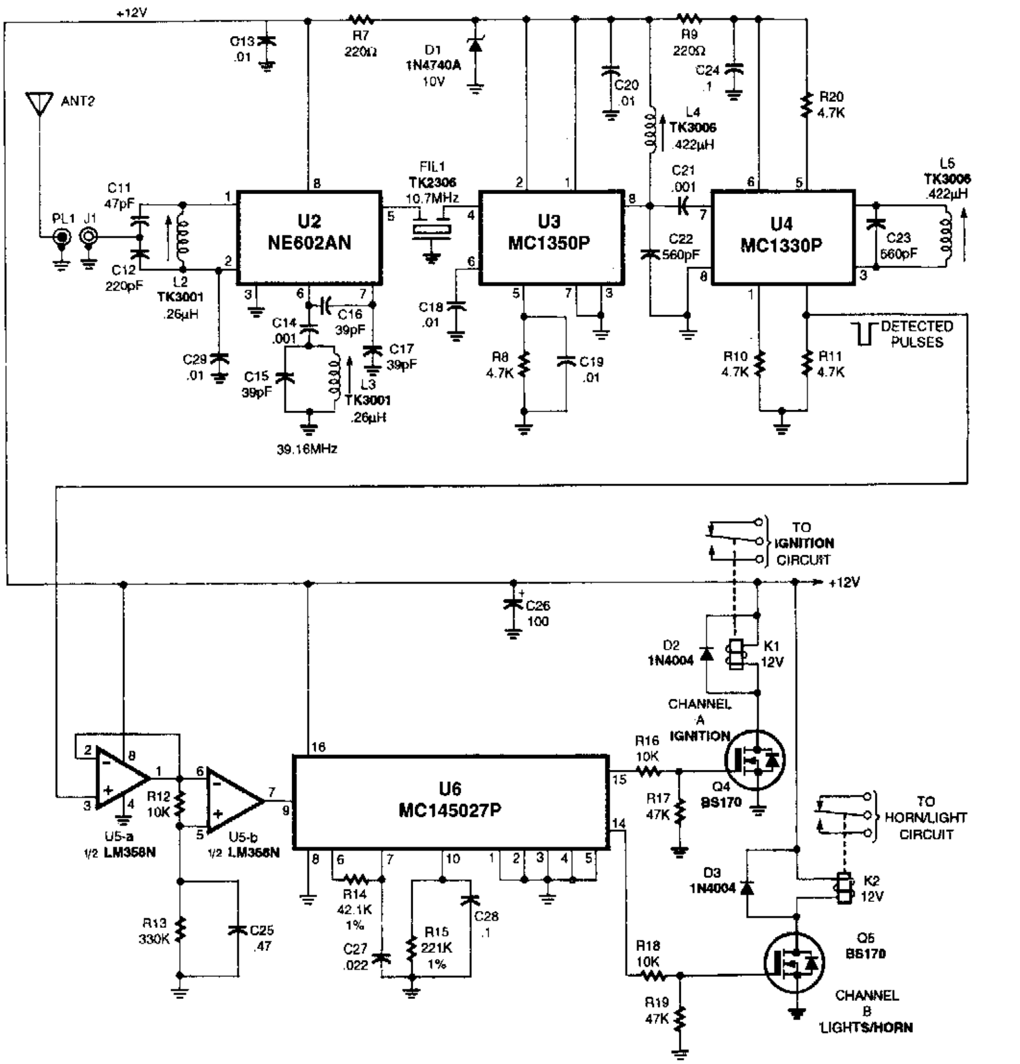
POPULAR ELECTRONICS

Fig. 3-3

At the heart of the main-control unit is U16, Zilog's Z8-8-bit microcontroller, which receives its program instruction from U17, a 27C64 8K x 8 EPROM. The home security system in its most basic form, features eight individual protection zones, adjustable entry and exit delays, a panic switch (for emergency situations), automatic system reset, support for an auto-dialer (which, in case of emergency, dials pre-programmed telephone numbers), it's X-10 compatible (allowing it to control house lights and appliances), has a backup battery (to keep the system on-line during a power failure), and there is also an optional zone-status panel that is used to individually show the condition of each protection zone.

This system's power supply provides 12 Vdc for the sirens and digital keypad, and 5 Vdc for the on-board electronics, while also providing a constant 12-V output that's used to charge the backup-battery.

AUTO SECURITY SYSTEM RECEIVER

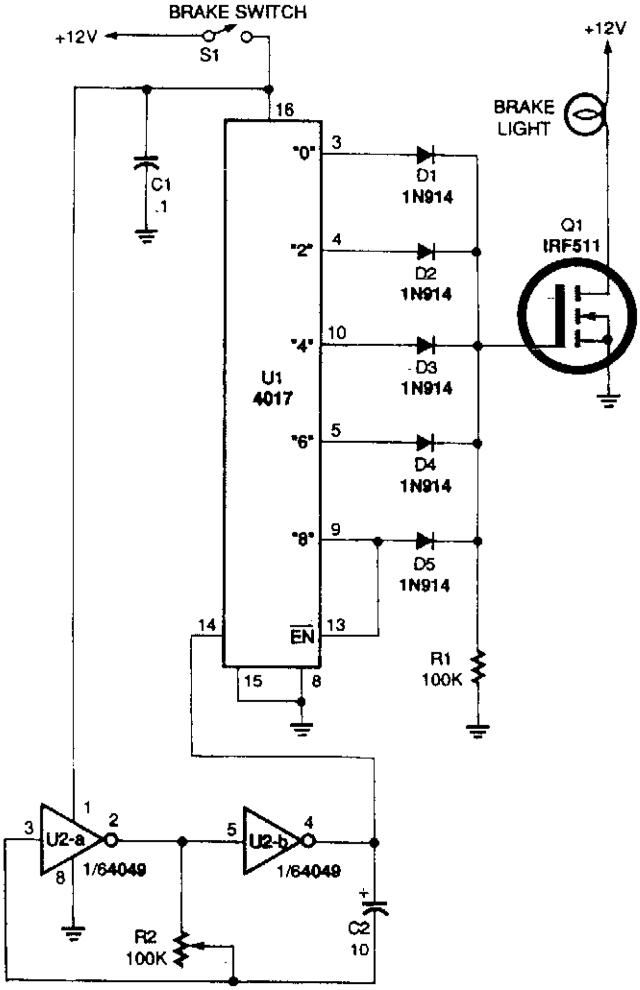


POPULAR ELECTRONICS

Fig. 3-4

This receiver is a superheterodyne type tuned to 49 MHz. U2 is a mixer, followed by a 10.7-MHz filter and two-stage IF (U3, U4) and detector. The encoded received RF pulse signal from the antenna produces detected pulse from the MC1330P. These pulses are amplified by U5 and fed to decoder IC U6, and MC1450278. Two channels are available at the output, which drives K1 and K2.

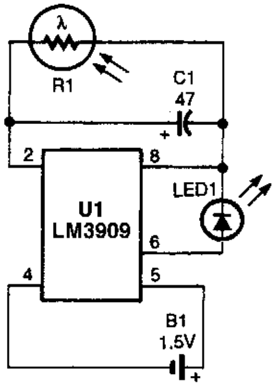
FLASHING BRAKE LIGHT FOR MOTORCYCLES



POPULAR ELECTRONICS *Fig. 3-5*

When brake-light switch S1 is closed, power is applied to U1 and U2. Two inverters of U2, a 4049 hex inverting buffer, are connected in a low-frequency oscillator circuit that feeds clock pulses into U1, a 4017 decade counter/divider. Outputs 0, 2, 4, 6, and 8 of U1 are coupled to the gate of Q1 through a 1N914 diode. As the 4017 counts down, it turns the brake light on and off four times and then leaves it on until the brake switch is released. The on/off rate can be set by potentiometer R2; for best results, the on/off rate should be set so that it is rapid.

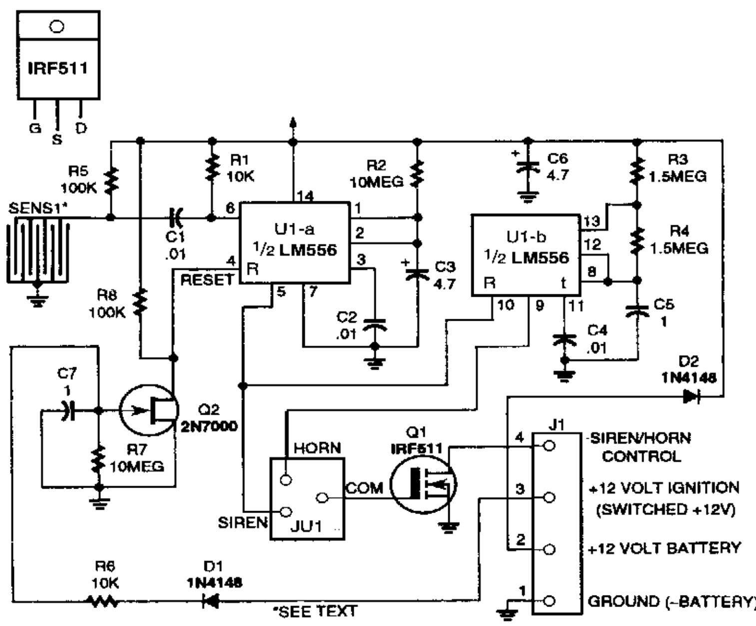
CAR ALARM DECOY



The device will simulate the presence of a burglar alarm in automobiles or homes. Mount R1 where daylight can fall on it. During darkness, LED1 flashes, making potential intruders think an alarm system is installed.

POPULAR ELECTRONICS *Fig. 3-6*

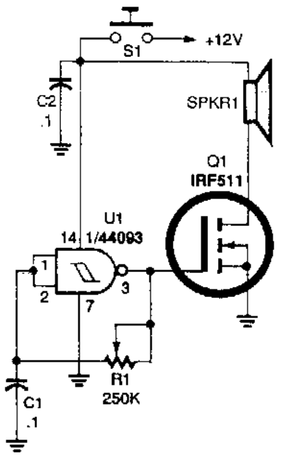
MOTORCYCLE ALARM



POPULAR ELECTRONICS *Fig. 3-7*

A dual timer is used to generate a long pulse, which gates a second timer, producing a square wave (nonsymmetrical) and controls the on/off time of the horn. Siren operation can be selected with a jumper. In this case, the output of Q1 will be continuously on and not cycled. Sensor S1 is a row of adjacent circuit board traces with a stainless steel ball bearing laying on them. Any movement causes momentary shorting and opening of the circuit, triggering U1-a.

SIMPLE BIKE HORN

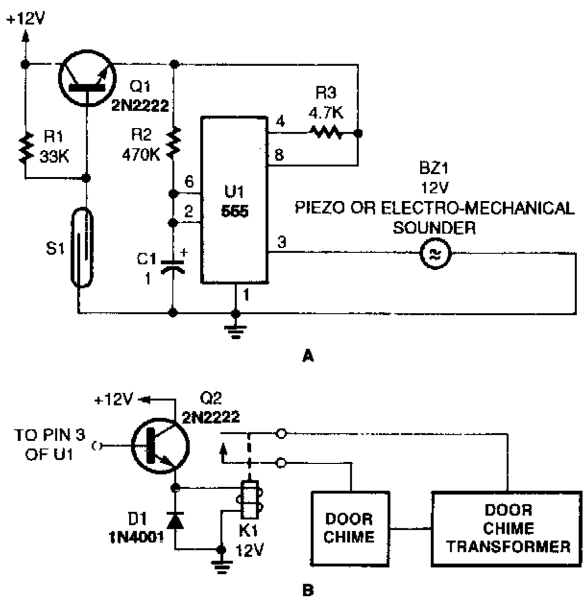


The horn circuit uses only one gate of a 4093 quad 2-input NAND Schmitt trigger, U1, connected in a simple, low-frequency, square-wave oscillator circuit. The oscillator's output, at pin 3, drives the gate of Q1. The drain of that FET drives a small horn speaker.

Potentiometer R1 can be adjusted to set the horn's output frequency. Some horn speakers are frequency sensitive, so play with the oscillator's frequency control for the best or loudest sound.

POPULAR ELECTRONICS *Fig. 3-8*

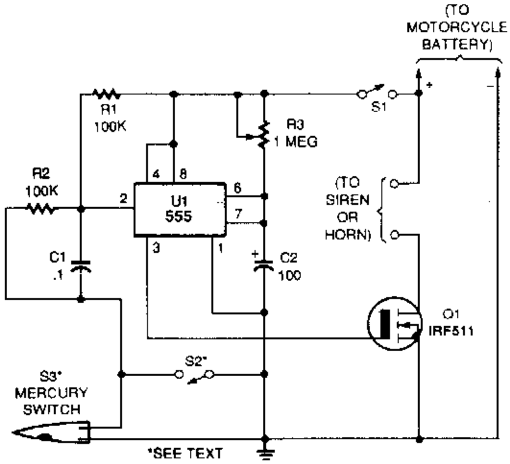
DOOR AJAR INDICATOR



POPULAR ELECTRONICS *Fig. 3-9*

This simple sounder (A) makes a good door annunciator. If the buzzer is replaced with the circuit in B, the annunciator can be made more pleasant to the ear.

MOTORCYCLE BURGLAR ALARM

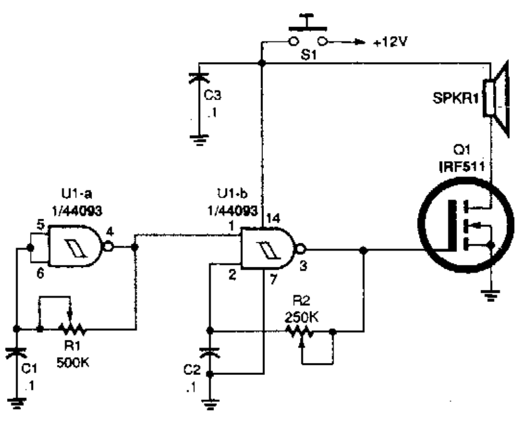


POPULAR ELECTRONICS *Fig. 3-10*

A 555 IC is connected in a one-shot timer circuit that turns on a FET transistor and either a siren or the bike's horn for a preset time period. Switch S1 is used as an on/off switch.

Closing either of two switches, S2 and S3, will trigger the IC. When either switch closes, pin 2 of U1 goes low. That triggers the IC to produce a positive output at pin 3 and sounds the alarm for the time period set by R3. The mercury switch, S3, is the switch that activates the alarm should anyone move your bike. Switch S2 can be used as a panic switch.

HORN CIRCUIT FOR MOTORCYCLE USE



POPULAR ELECTRONICS *Fig. 3-11*

Gates U1-a and U1-b of the 4093 quad 2-input NAND Schmitt trigger are connected in variable, low-frequency, square-wave oscillator circuits. The output of gate U1-a is connected to one of the inputs of gate U1-b. The square-wave output of gate U1-a modulates oscillator U1-b, producing a two-tone output. A really interesting sound can be produced by carefully adjusting potentiometers R1 and R2.

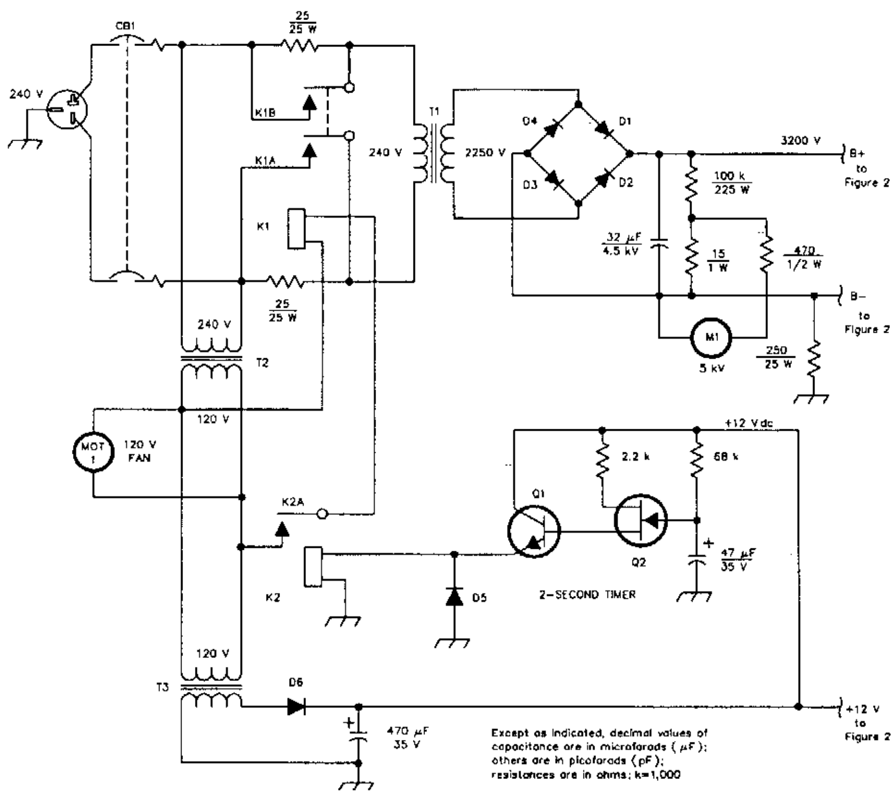
4

Amateur Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

1.2-kW 144-MHz Amplifier Power Supply
1.2-kW 144-MHz Amplifier Control Circuitry
1.2-kW 144-MHz Linear Amplifier
Four-Stage 75-Meter SSB Superhet Receiver
Improved CW Transmitter Keying Circuit
One-Chip AFSK Generator
Programmable CW Identifier
Audible SWR Detector Adapter
Audio Breakout Box
One-Watt CW Transmitter
PTT Control from Receiver Audio
Transceiver Memory Backup
80-Meter SSB Receiver
CW Audio Filter
RF Line Sampler/Coupler
Battery Pack and Reverse Polarity Protection
Simple Identifier
Transmit Keyer Interface Circuits
Mobile Radio On-Alarm Timer

1.2-kW 144-MHz AMPLIFIER POWER SUPPLY



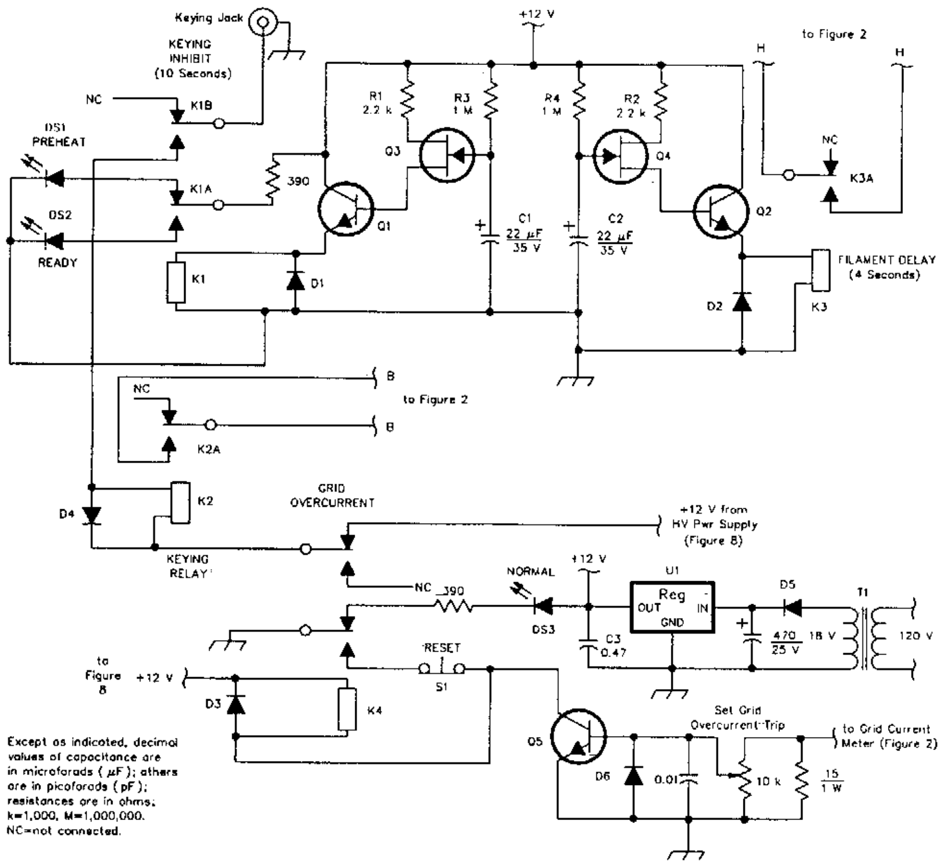
Schematic diagram of the high-voltage power supply recommended for use with the power amplifier.

D1-D4—Strings of 4 each, 1000-PIV, 3-A diodes, 1N5408 or equivalent
K1—DPST relay, 120-V ac coil, 240-V-ac, 20-A contacts (Midland Ross 187-321200 or equivalent)
K2—SPDT miniature relay, 12-V dc coil (Radio Shack 275-248 or equivalent)
M1—High-voltage meter, 5 kV dc full scale (1-mA meter movement used with series resistors shown in drawing)
MOT1—Cooling fan, Torin TA-300 or equivalent
Q1—2N2222A or equivalent
Q2—MPF102 or equivalent
S1—20-A hydraulic/magnetic circuit breaker (Potter and Brumfield W68X2Q12-20 or equivalent)
T1—High-voltage power transformer, 240-V primary, 2250-V, 1.2-A secondary (Avatar AV-538 or equivalent)
T2—Stepdown transformer, Jameco 112125, 240-V to 120-V, 100 VA
T3—Power transformer, Jameco 104379, 120-V primary; 16.4 V, 1-A secondary (half used)

QST *Fig. 4-1*

A schematic diagram of the high-voltage power supply recommended for use with the power transformer. This power supply can also be used for other equipment with similar requirements. CAUTION: hazardous high voltages.

1.2-kW 144-MHz AMPLIFIER CONTROL CIRCUITRY

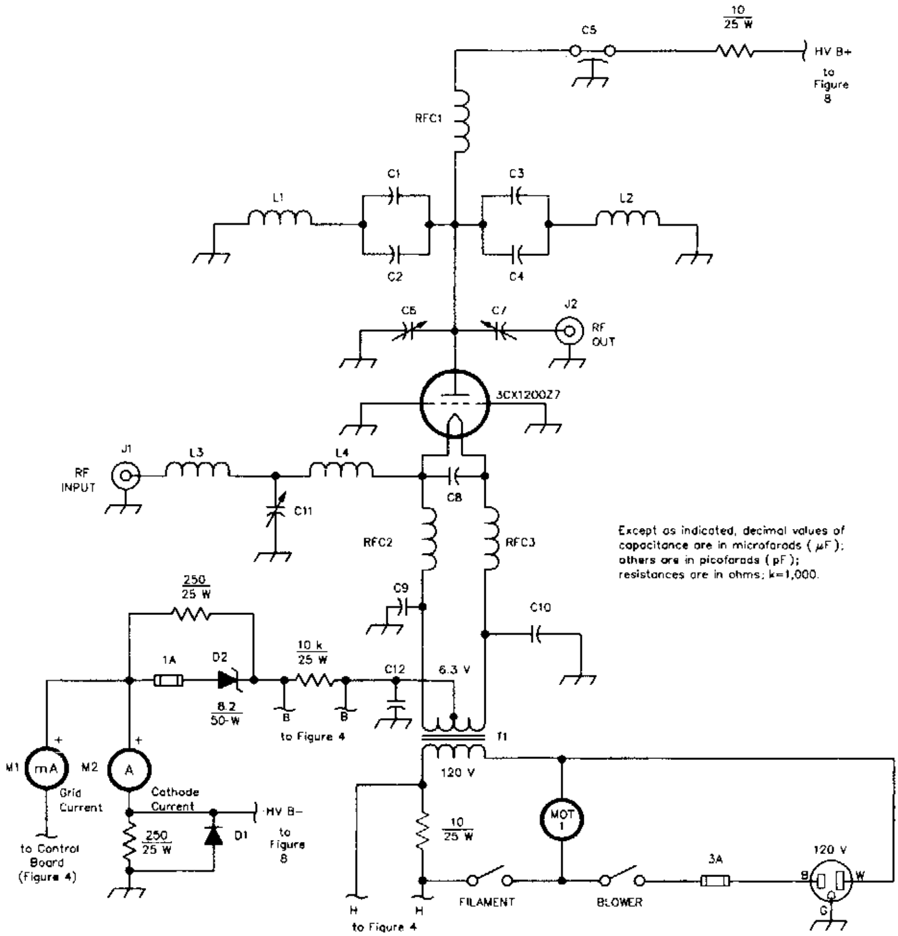


QST

Fig. 4-2

Schematic diagram of the amplifier-control circuits.

1.2-kW 144-MHz LINEAR AMPLIFIER



C1-C4—100 pF, 5 kV, type 850
C5—1000 pF, 5 kV
C6—Anode-tuning capacitor; see text and Figure 5 for details
C7—Output-loading capacitor; see text and Figure 7 for details
C8-C10—1000-pF silver mica, 500 V
C11—30-pF air variable
C12—0.01 µF, 1 kV
D1—1000 PIV, 3-A diode, 1N5408 or equiv
D2—8.2-V, 50-W Zener diode, ECG 5249A
J1—Chassis-mount BNC connector
J2—Type-N connector fitted to output coupling assembly (see Figure 7)
L1, L2—Plate lines; see text and Figure 6 for details
L3—5 t no. 14, 1/2-inch diameter, close wound
L4—3 t no. 14, 5/8-inch diameter, 1/4-inch spacing
RFC1—7 t no.14, 5/8-inch diameter, 1 3/8 inch long
RFC2, RFC3—10 t no. 12, 5/8-inch diameter, 2 inches long
T1—Filament transformer. Primary: 120 V; secondary: 6.3 V, 25 A, center tapped Available from Avatar Magnetics (Ronald C. Williams, W9JVF, 240 Tamara Trail, Indianapolis, IN 46217, 317-783-1211); part number AV-539
M1—Grid milliammeter, 200 mA dc full scale
M2—Cathode ammeter, 2 A dc full scale
MOT1—140 free-air cfm, 120-V ac blower, Dayton 4C442 or equivalent.

Sources for some of the "hard to get parts" include:

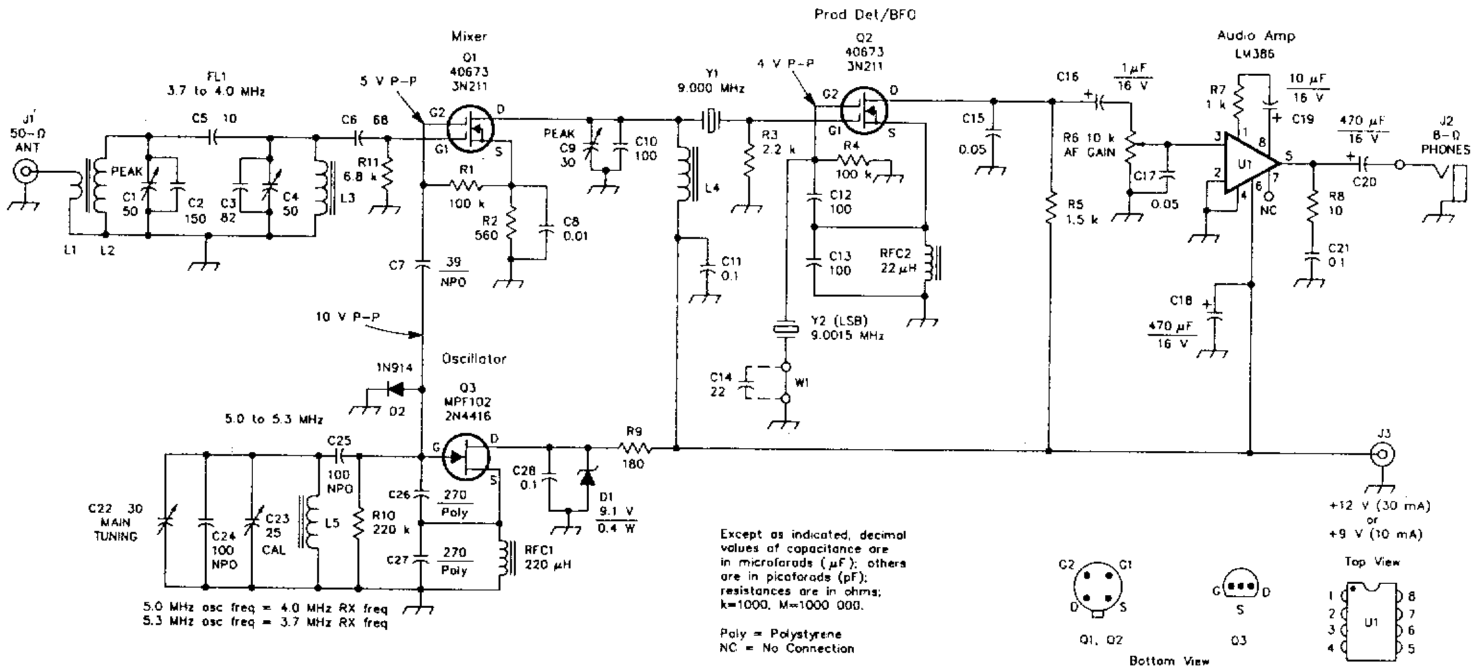
Fair Radio Sales, 1016 E Eureka, Lima, OH 45802, tel 419-227-6573

Surplus Sales of Nebraska, 1502 Jones Street, Omaha, NE 68102, tel 402-346-4750.

QST **Fig. 4-3**

Schematic diagram of the 2-meter amplifier.

FOUR-STAGE 75-METER SSB SUPERHET RECEIVER

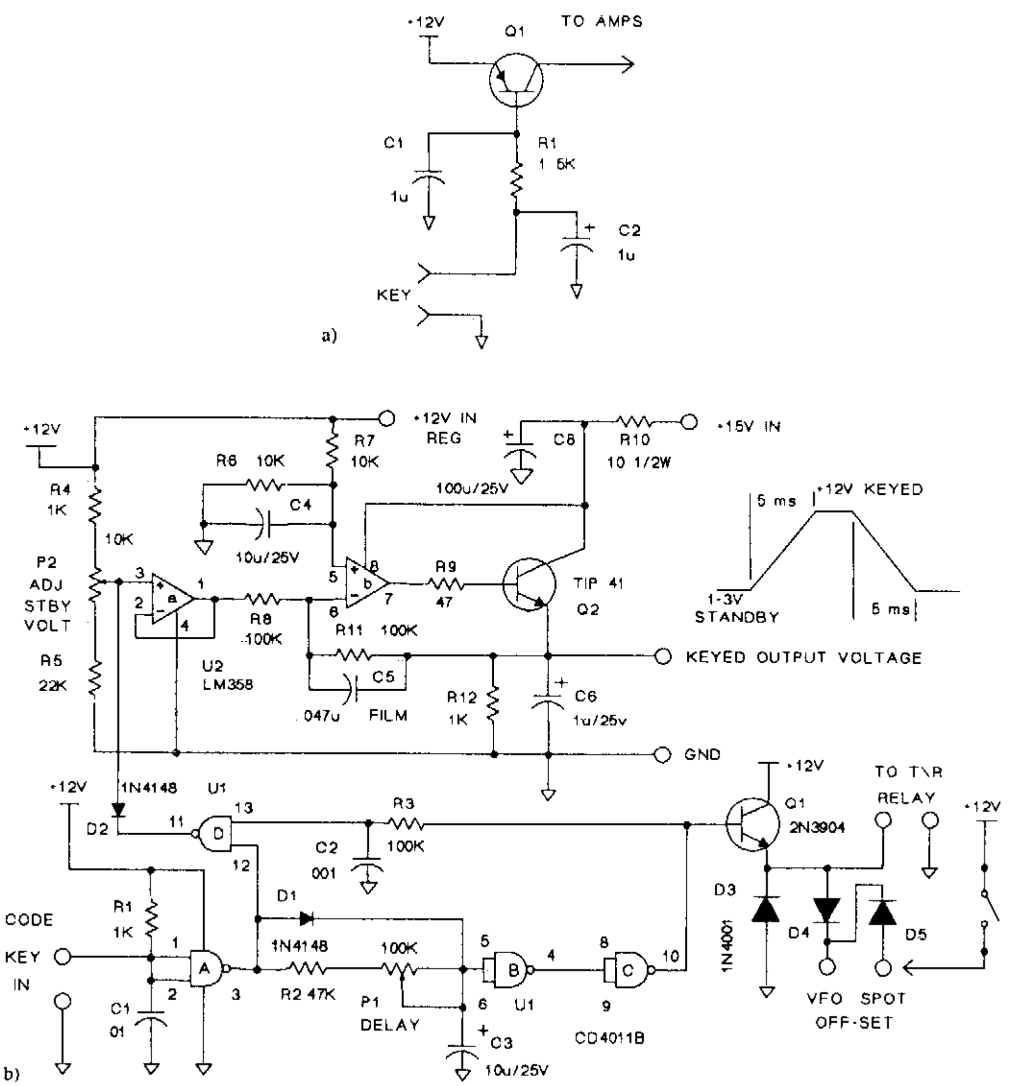


QST

Fig. 4-4

A simple superhet receiver for SSB reception in the 75-meter amateur band is shown. Y1 acts as a crystal filter.

IMPROVED CW TRANSMITTER KEYING CIRCUIT



73 AMATEUR RADIO TODAY ***Fig. 4-5***

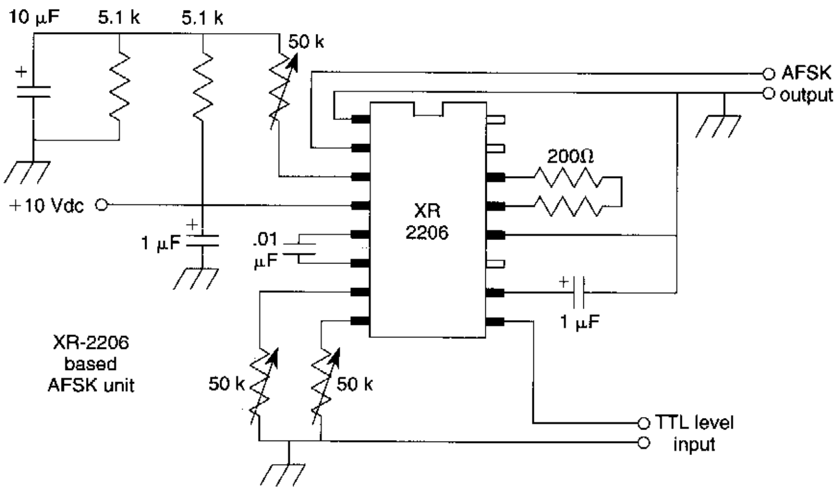
Typical (A) QRP keying circuit; (B) Improved QRP keying circuit with CMOS T/R keying control. Op amp U2b is a basic inverting amplifier with a gain of one. The capacitor C5 across the feedback resistor R11 makes it an integrator. The RC time constant of R11 and C5 determine the ramp time. The values shown will produce a 5-ms ramp. Use a good-quality capacitor for C5, such as a mylar or polypropylene type. A power transistor is placed inside the feedback loop so that the circuit can supply several hundred milliamperes of current. Control P2 sets the stand-by output voltage as seen at the emitter of Q1. U2a buffers the voltage from P2. This isolates the pot from the input of the integrator. With the key up, adjust the pot until you just start to see an output from your transmitter, then back it off a little. Typically, this will be between 2 and 4 V. Your output signal will now have the proper 5-ms leading and falling edges and there will be no delay between key closure and the start of the output signal.

You must supply the op amp and collector of Q2 with at least 15 V to produce a full 12-V output on the emitter.

Parts list

R1, R4, R12	1 kΩ ¼ W	C1	0.01 µF disk
R2	47 kΩ ¼ W	C2	0.001 µF disk
R3, R8, R11	100 kΩ	C3, C4	10 µF, 25 V electrolytic
R6, R7	22 kΩ	C5	0.047 µF poly-film type
R9	47 Ω	C6	1 µF, 25 V electrolytic
R10	10 Ω	C7	Skipped
P1	100 kΩ or 500 kΩ trimpot	C8	100 µF, 26 V electrolytic
P2	10 kΩ trimpot	D1, D2	1N4148 diode
		D3, D4, D5	1N4001 1-A diode
Q1	2N3904 NPN		
Q2	Tip 41-to-220 NPN		
U1	4011B CMOS NAND gates		
U2	LM358 dual op amp		

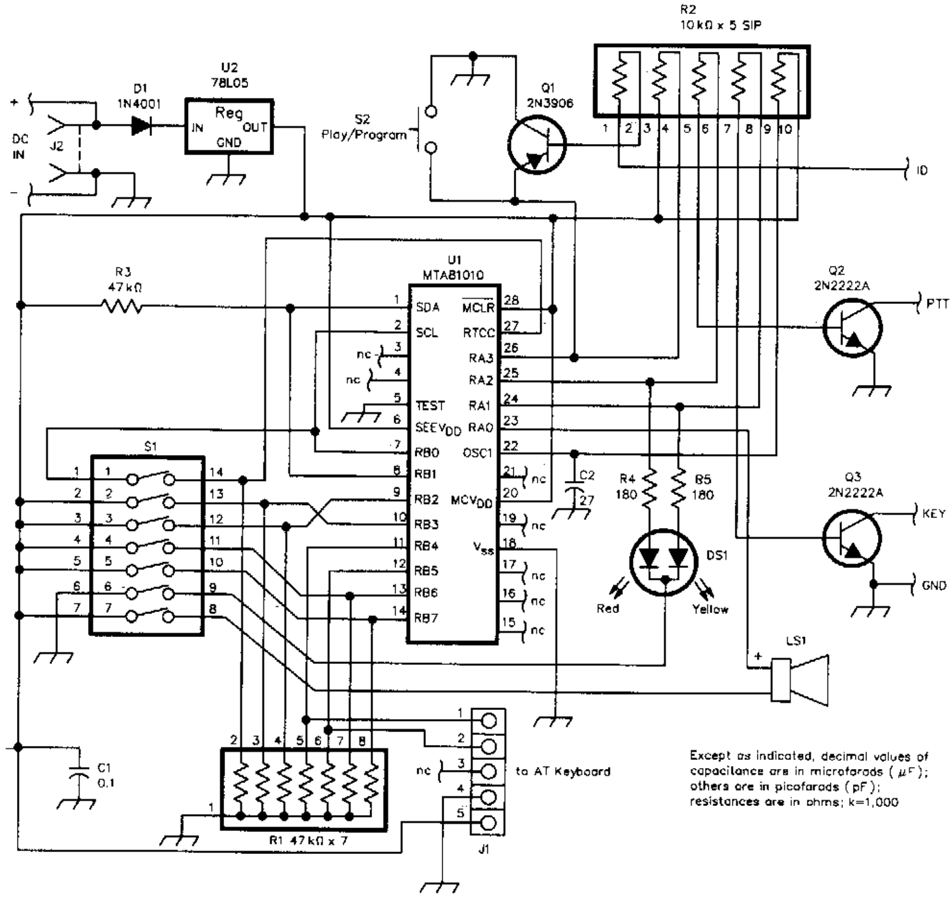
ONE-CHIP AFSK GENERATOR



73 AMATEUR RADIO TODAY *Fig. 4-6*

Built around an XR2206 IC, this circuit will generate AFSK signals in the 1000- to 3000-Hz range.

PROGRAMMABLE CW IDENTIFIER



S1 Control Settings

WPM	*5*	*7.5*	*10*	*13*	*15*	*18*	*20*	*25*
S1-1	off	off	off	off	on	on	on	on
S1-2	off	off	on	on	off	off	on	on
S1-3	off	on	off	on	off	on	off	on

TIMER	*OFF*	*5 Min*	*10 Min*	*Continuous*
S1-4	off	off	on	on
S1-5	off	on	off	on

S1-6—DS1 on/off
S1-7—Speaker on/off

PC-Board Connections

ID—A momentary ground on this terminal causes the IDer to play its message; same as pressing the **PLAY/PROGRAM** pushbutton.

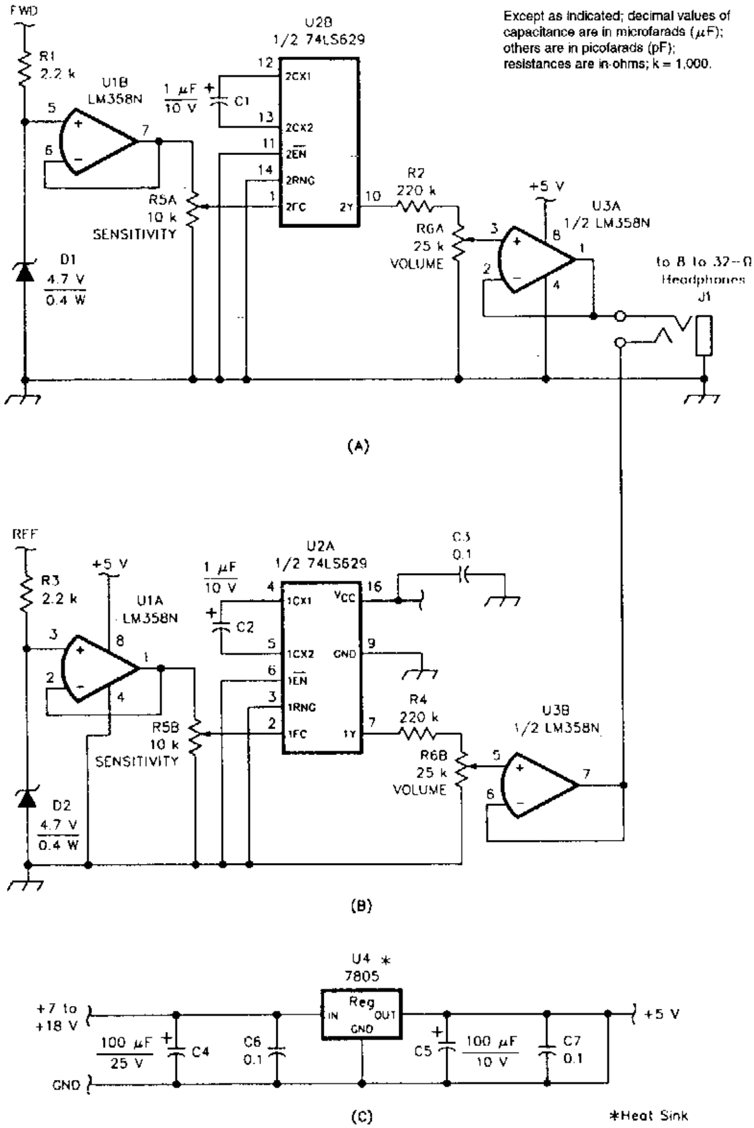
PTT—An open-collector output which goes to ground 250 ms before the CW output occurs. This output is used to place radio in transmit mode and is monitored by the red LED.

KEY—An open-collector output that goes to ground during CW keying. This output is monitored by the speaker and the yellow LED.

QST *Fig. 4-7*

The identifier uses an MTA81010 microchip, containing a 1024-bit serial EEPROM and a microcontroller. It runs from a 9-V battery. A standard AT-type keyboard is used to program the desired message. Speed varies from 5 to 25 wpm.

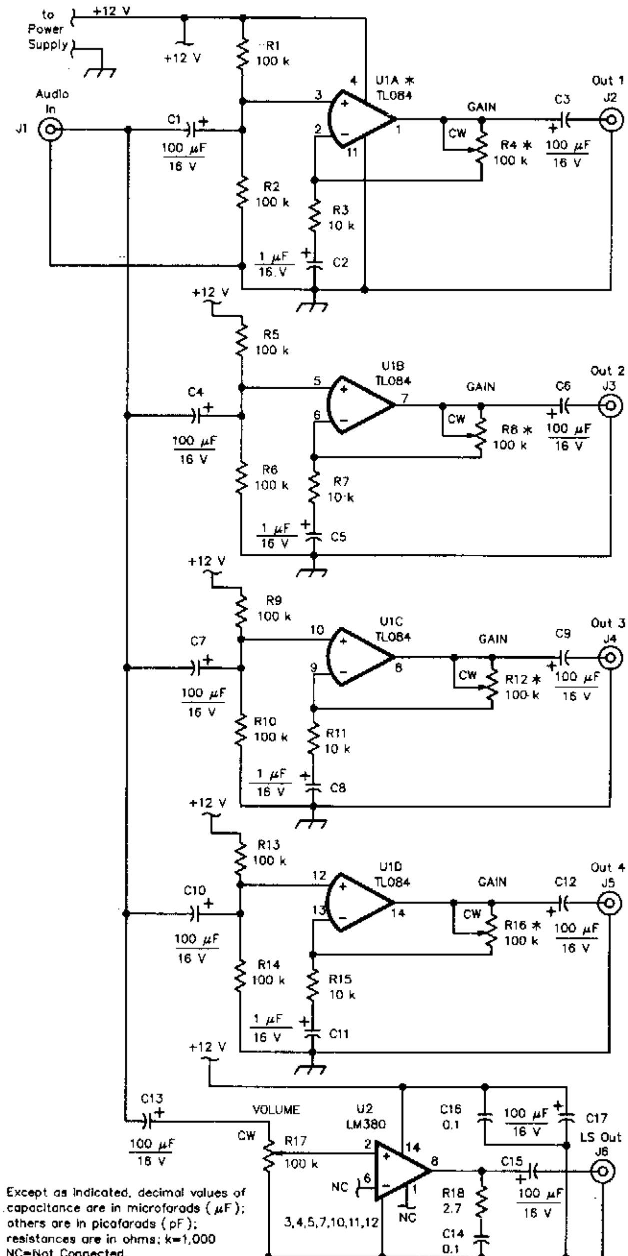
AUDIBLE SWR DETECTOR ADAPTER



QST *Fig. 4-8*

This SWR detector audio adapter is designed specifically for blind or vision-impaired amateurs, but anyone can use it. Instead of using a meter (or meters) to indicate antenna system forward and reflected voltages, this adapter generates two tones with frequencies that are proportional to the respective voltages. The tones are fed to a pair of stereo headphones (the miniature types are ideal) so that one ear hears the forward-voltage tone and the other ear hears the reflected-voltage tone. Thus, tuning up a transmitter is simply a matter of tuning for the highest-pitched tone in the left ear and the lowest-pitched tone in the right ear.

AUDIO BREAKOUT BOX



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000
NC=Not Connected.
IC pins not shown are unused.
∗=See text.

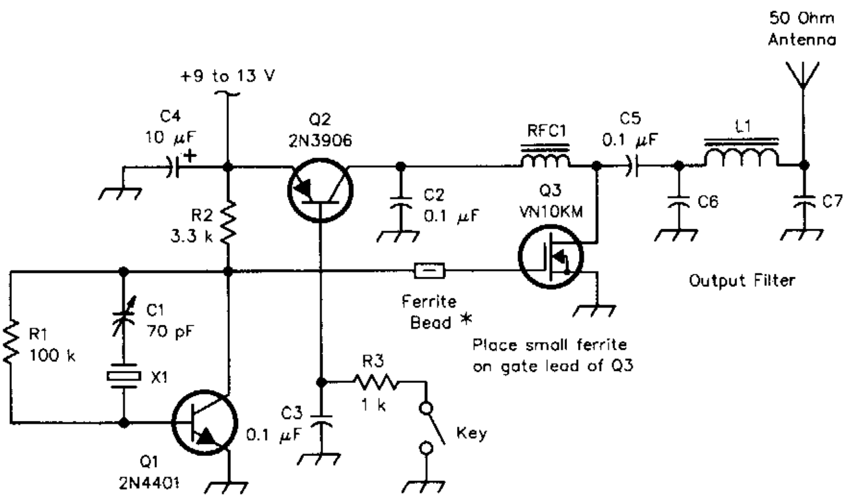
QST *Fig. 4-9*

AUDIO BREAKOUT BOX (*Cont.*)

In many radio shacks, one receiver audio-output line feeds a multitude of add-ons, such as one or more TNXs, SSTV modems, PC plug-in boards, and, perhaps, speakers. Having to manually plug the audio source from one accessory to another is inconvenient, if not frustrating as well. Overloading the sources by connecting the loads in parallel isn't satisfactory, either.

The audio breakout box takes the audio output from a receiver (or other audio source) and applies it to the inputs of four identical, independent, low-level AF buffer/amplifiers and one high-level (1-W output) AF channel. Each low-level output channel can provide up to 20 dB of gain that's independently adjustable.

ONE-WATT CW TRANSMITTER



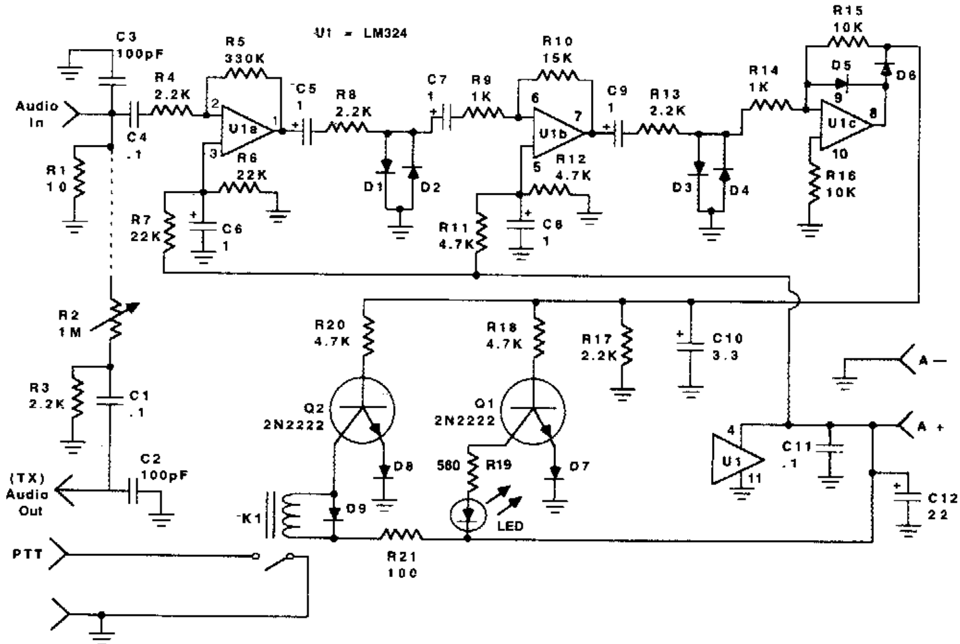
QST

Fig. 4-10

C6,C7
820 pF disc ceramic (160 meters)
470 pF disc ceramic (80 meters)
220 pF disc ceramic (40 meters)
150 pF disc ceramic (30 meters)
100 pF disc ceramic (20 meters)
82 pF disc ceramic (17 meters)

L1
33 turns, #30, T37-2 (160 meters)
23 turns, #30, T37-2 (80 meters)
17 turns, #26, T37-2 (40 meters)
14 turns, #26, T37-2 (30 meters)
12 turns, #26, T37-2 (20 meters)
10 turns, #26, T37-2 (17 meters)

PTT CONTROL FROM RECEIVER AUDIO

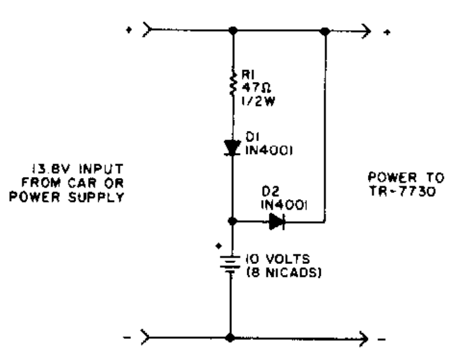


73 AMATEUR RADIO TODAY

Fig. 4-11

This system will key a transmitter or other device that can be keyed by a relay closure. Audio is amplified, limited, and is rectified and drives relay driver Q2 and LED indicator. The transmitter audio output was used to feed a keyed transmitter and can be deleted or ignored where this feature is unnecessary.

TRANSCEIVER MEMORY BACKUP

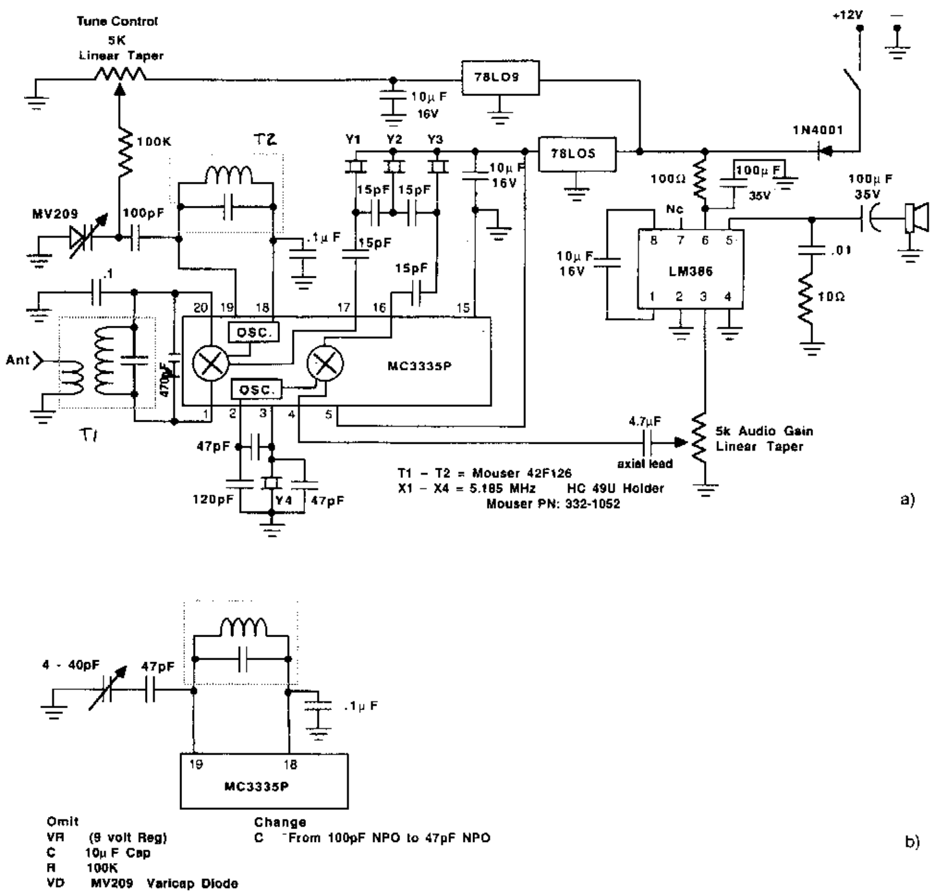


73 AMATEUR RADIO TODAY

Fig. 4-12

Although designed for a Kenwood TR7730, this idea might be adapted to other transceivers. This circuit will retain the frequencies in memory while moving the rig from car to house and vice versa. When connected to an external power source, battery B1 is charged through R1 and D1. D1 prevents B1 from discharging when connected to an external supply that is turned off. When external power is removed, D2 provides a current path to the TR-7730 to retain the memory's contents. However, the TR-7730 power switch should be turned off before external power is removed because B1 will not provide power for normal operation.

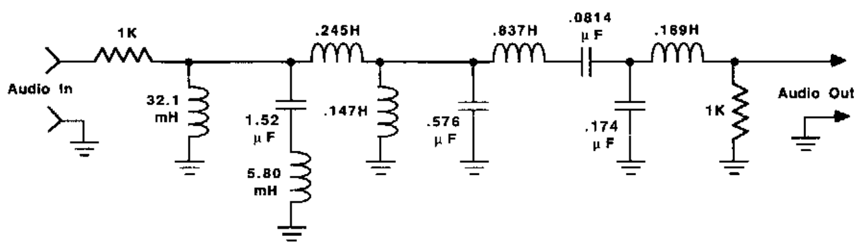
80-METER SSB RECEIVER



73 AMATEUR RADIO TODAY *Fig. 4-13*

This circuit uses an MC3335P IF chip and features a 3-pole crystal filter made from microprocessor crystals. Tuning is done either with a varactor diode or air-variable capacitor, as shown. Values are for 80 meters.

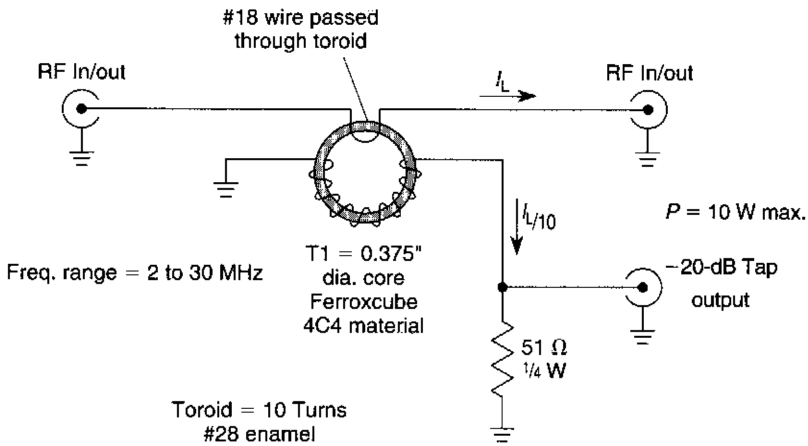
CW AUDIO FILTER



73 AMATEUR RADIO TODAY *Fig. 4-14*

A high-performance passive filter. The center frequency is 700 Hz; –3-dB bandwidth is 200 Hz.

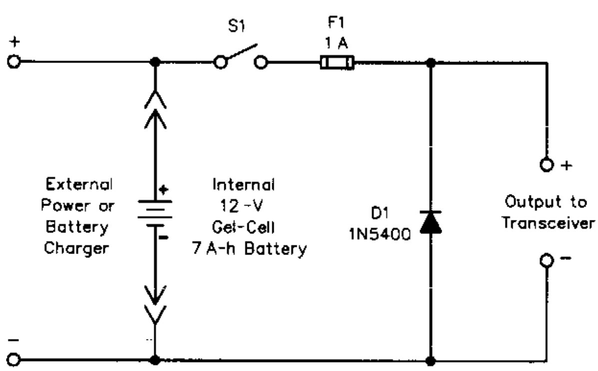
RF LINE SAMPLER/COUPLER



WILLIAM SHEETS *Fig. 4-15*

Suitable for RF line sampling, this coupler is useful where an indirect measurement of line current is needed. A 10:1 turn ratio yields a secondary current about 1/10 (ideally) of the line current. A 51-Ω resistor terminates the secondary. Insertion loss in the main line is negligible, < 0.1 dB. For higher power levels, use proportionately larger core for T1.

BATTERY PACK AND REVERSE POLARITY PROTECTION

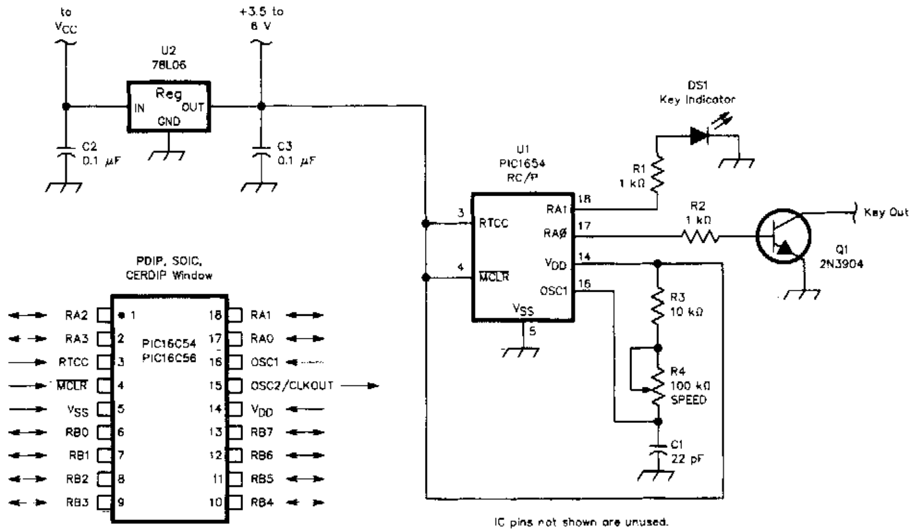


QST *Fig. 4-16*

Schematic diagram and parts list for the reverse-polarity protection circuit (optional).

D1 1N5400 silicon diode
F1 1-A fast-acting fuse
S1 SPST rocker switch

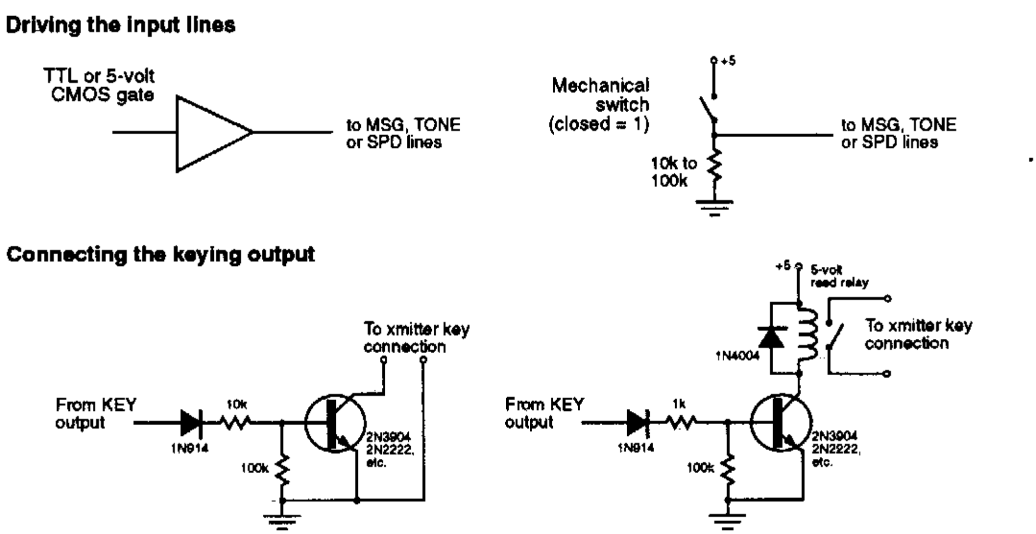
SIMPLE IDENTIFIER



QST *Fig. 4-17*

This identifier uses a PIC 16C54 microcontroller which must be programmed for your desired identifier.

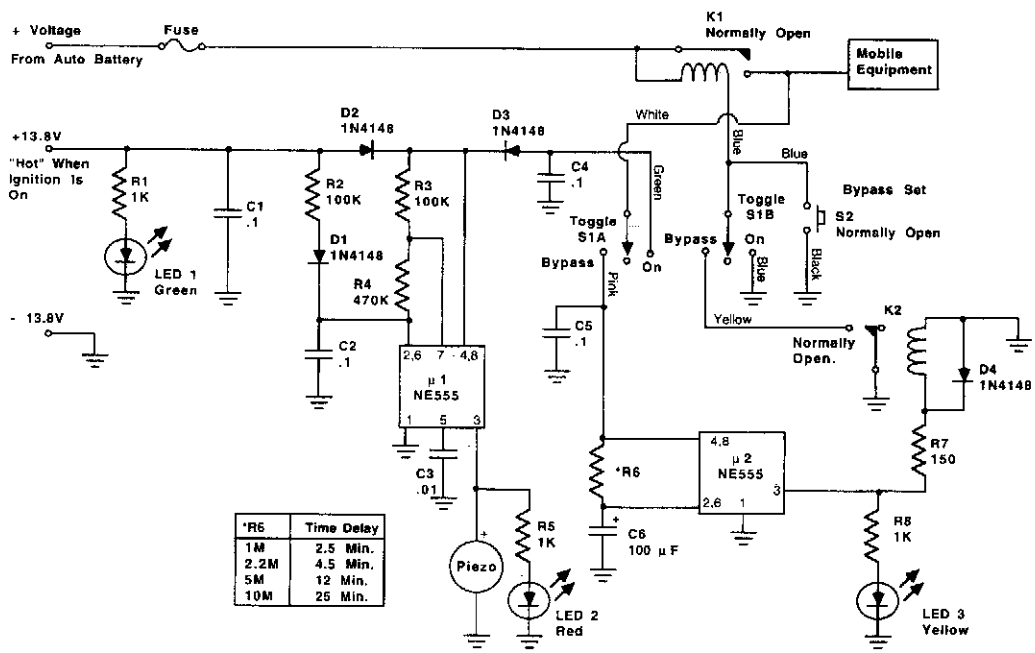
TRANSMIT KEYER INTERFACE CIRCUITS



73 AMATEUR RADIO TODAY *Fig. 4-18*

These circuits are for use with Morse keyers and identifiers. They can be used to interface various devices with the identifier circuitry.

MOBILE RADIO ON-ALARM TIMER



*R6	Time Delay
1M	2.5 Min.
2.2M	4.5 Min.
5M	12 Min.
10M	25 Min.

73 AMATEUR RADIO TODAY ***Fig. 4-19***

This system will prevent you from accidentally leaving your mobile radio on, draining the battery. LED1 will light when the vehicle's ignition is on, or while the car is running. Switch S1 in the ON position will close relay K1, completing the power circuit to the equipment. If the ignition switch is shut off, and switch S1 is still in the ON position, an alarm (piezo) will begin to beep and LED2 will flash. Returning S1 to the center position will shut everything off. If equipment operation is desired after shutting off the vehicle, you can place switch S1 in the AUTO position and momentarily press S2, a normally open push-button switch. Depressing this switch begins a timing cycle. The length of time that the Mobile-ON alarm/timer operates before shutting everything off can be "programmed" by selecting R6. The approximate time delays are provided in the chart with the schematic. Or, you could change the value of C6. These components control the holding time of relay K1. LED3 will light while the circuit is in AUTO status. Incidentally, you can also cancel the time delay at any time during the delay period by simply switching it off.

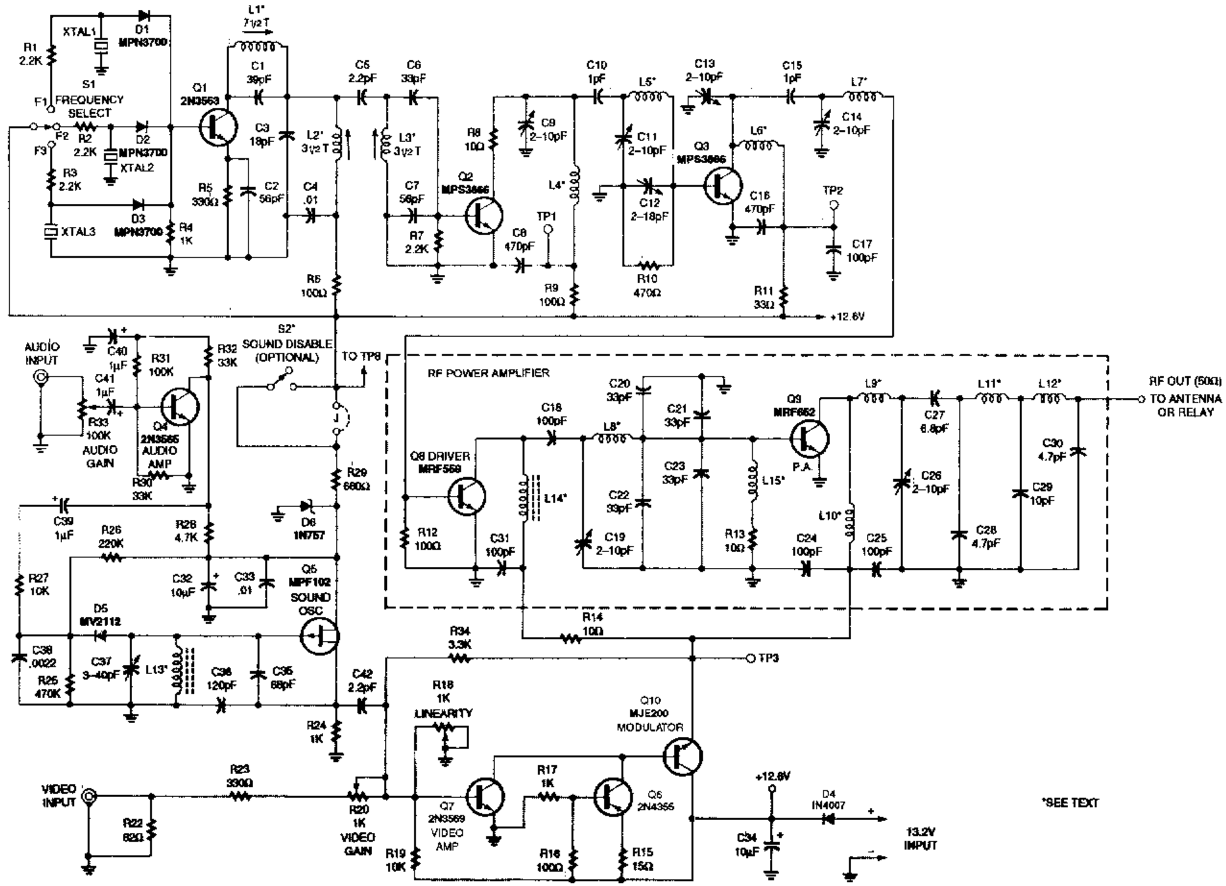
5

Amateur Television (ATV) Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

5-W ATV Transmitter for 440 MHz
5-W ATV Transceiver
Mini ATV Transmitter
Dummy Load and Video Detector for Transmitter Tests
Mast-Mounted ATV Preamp
Three-Channel 902- to 928-MHz ATV Transmitter
ATV Downconverter for 902 to 928 MHz
Three-Channel 420- to 450-MHz ATV Transmitter
ATV Downconverter for 420 to 450 MHz

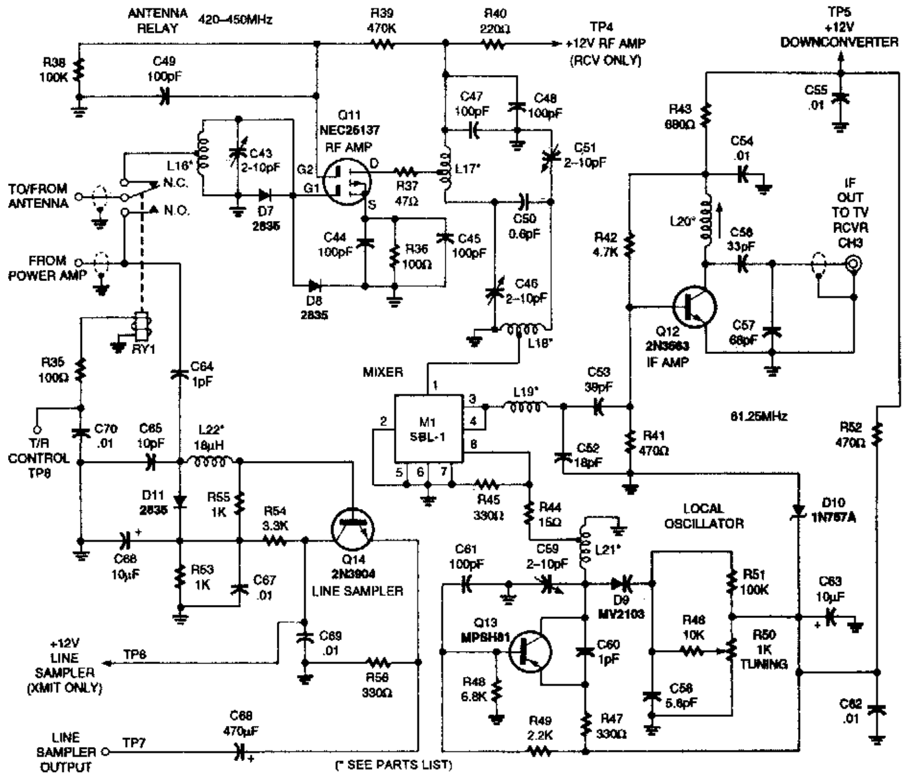
5-W ATV TRANSMITTER FOR 440 MHz



ELECTRONICS NOW *Fig. 5-1*

The circuit will produce typically 6 W RF output on synch tips. A crystal oscillator drives a doubler to produce a 220-MHz output. Another doubler produces 440 MHz to drive the power amplifier. A high-level series modulator provides the video modulation capability. A sound subcarrier is generated using a VCO circuit and combined with the video information. A complete kit of parts, including the PC board, is available from North Country Radio, P.O.Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

5-W ATV TRANSCEIVER

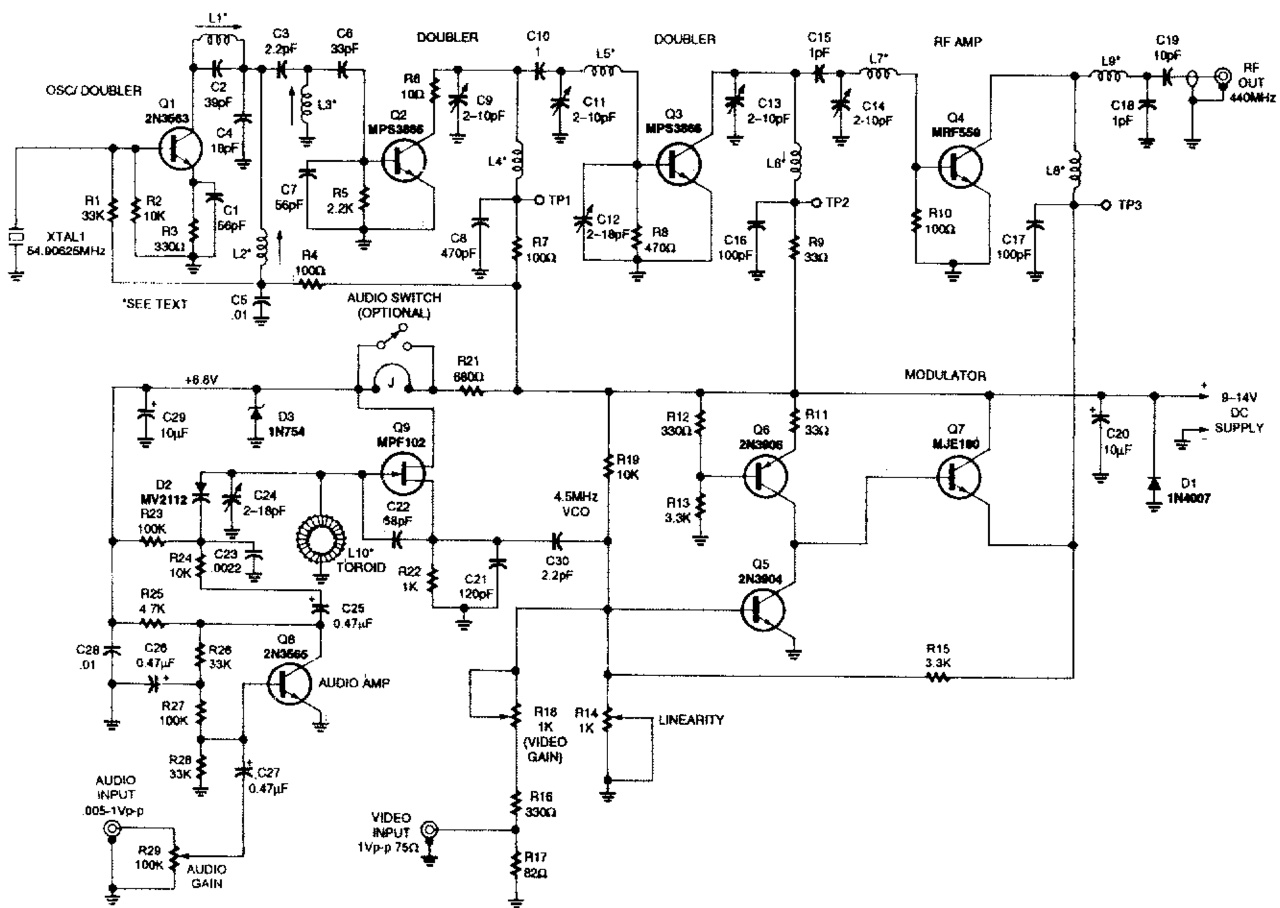


ELECTRONICS NOW *Fig. 5-2*

For the transmitter schematic (part of this transceiver), see entry entitled "5-W ATV Transmitter for 440 MHz, Fig. 5-1." The downconverter portion is shown here.

This transmitter contains both a video and sound section. Five to six watts PEP on synch tips of NTSC video are produced. Three channels are available. Channel switching is via PIN diodes. Power supply voltage is 12 to 14 Vdc. The receiver function is provided with a downconverter circuit and is tunable. A relay is used for T-R switching. A complete kit of parts, including PC board, is available from North Country Radio, P.O.Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

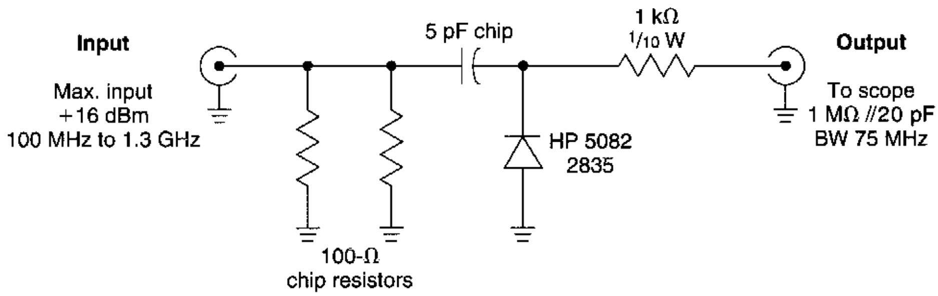
MINI ATV TRANSMITTER



ELECTRONICS NOW *Fig. 5-3*

This low-power NTSC video and sound transmitter is useful for amateur radio, video handie-talkie, R/C and surveillance purposes. A crystal oscillator-multiplier RF power amplifier. Video modulation is via a three-transistor series modulator. The sound subcarrier is generated with a VCO circuit and is combined with the video information. The output is 0.4 to 1.2 W with supply voltages of 9 to 14 volts. A complete kit of parts, including PC board, is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

DUMMY LOAD AND VIDEO DETECTOR FOR TRANSMITTER TESTS

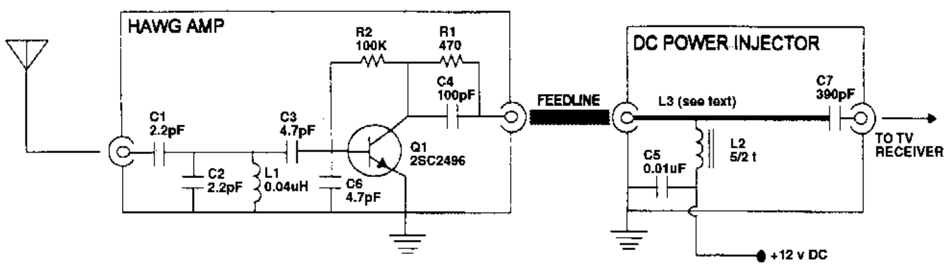


WILLIAM SHEETS

Fig. 5-4

This circuit is useful as a video modulation monitor for testing low-power video transmitters. For higher power inputs, use a suitable attenuator between the detector and the source. The detector should be connected to scope with as short a cable as possible to preserve video bandwidth.

MAST-MOUNTED ATV PREAMP

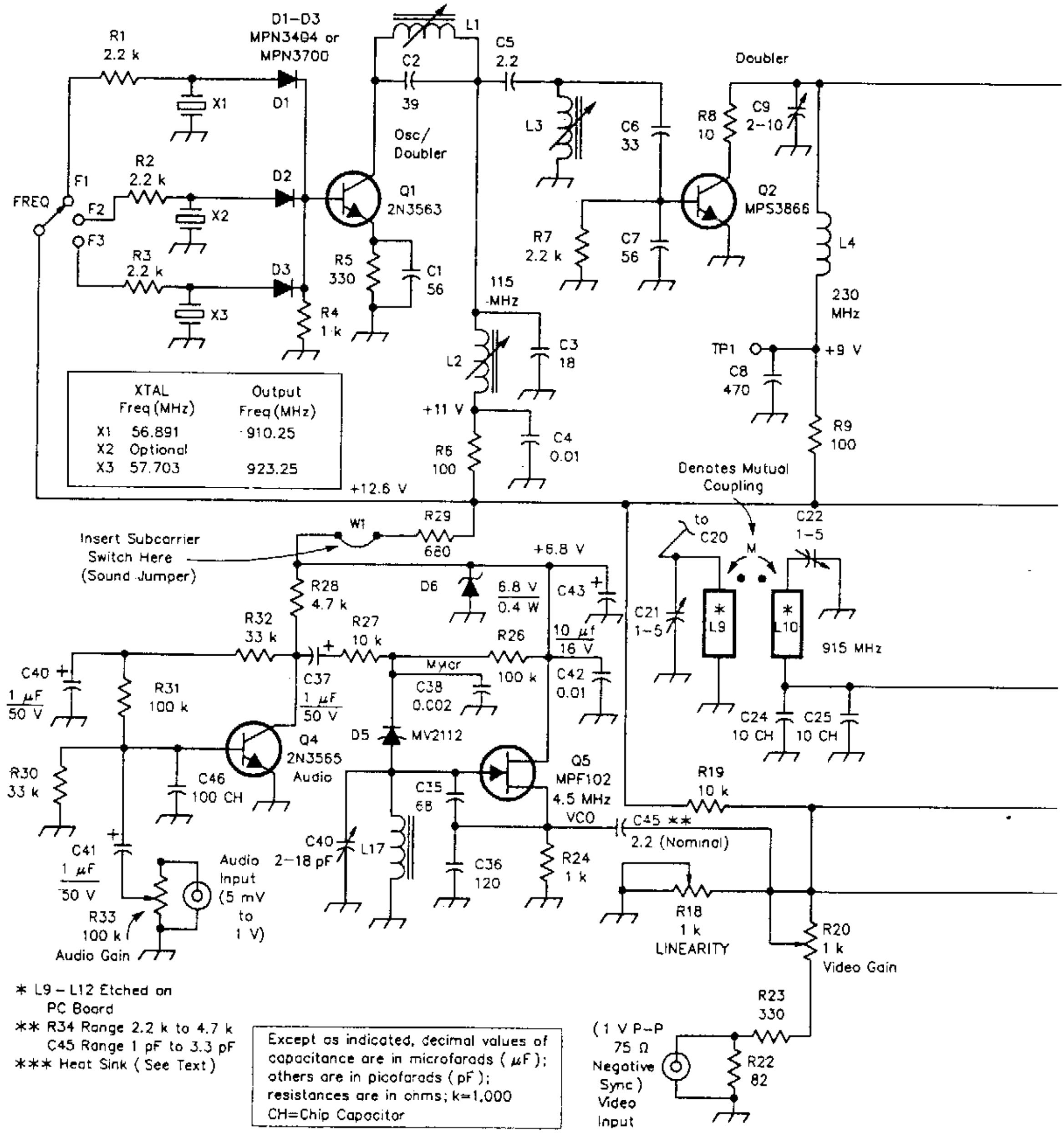


73 AMATEUR RADIO TODAY

Fig. 5-5

This simple ATV preamp covers the 427- to 439-MHz ATV frequencies and can be mast mounted and dc powered through the feedline.

THREE-CHANNEL 902- TO 928-MHz ATV TRANSMITTER



QST

This transmitter is for ATV applications in the 902- to 928-MHz band. It has three crystal-controlled channels, and will accept standard NTSC video input. It also has a 4.5-MHz sound subcarrier. Because this is an AM transmitter, audio can be transmitted as AM on the RF carrier. Simply use the

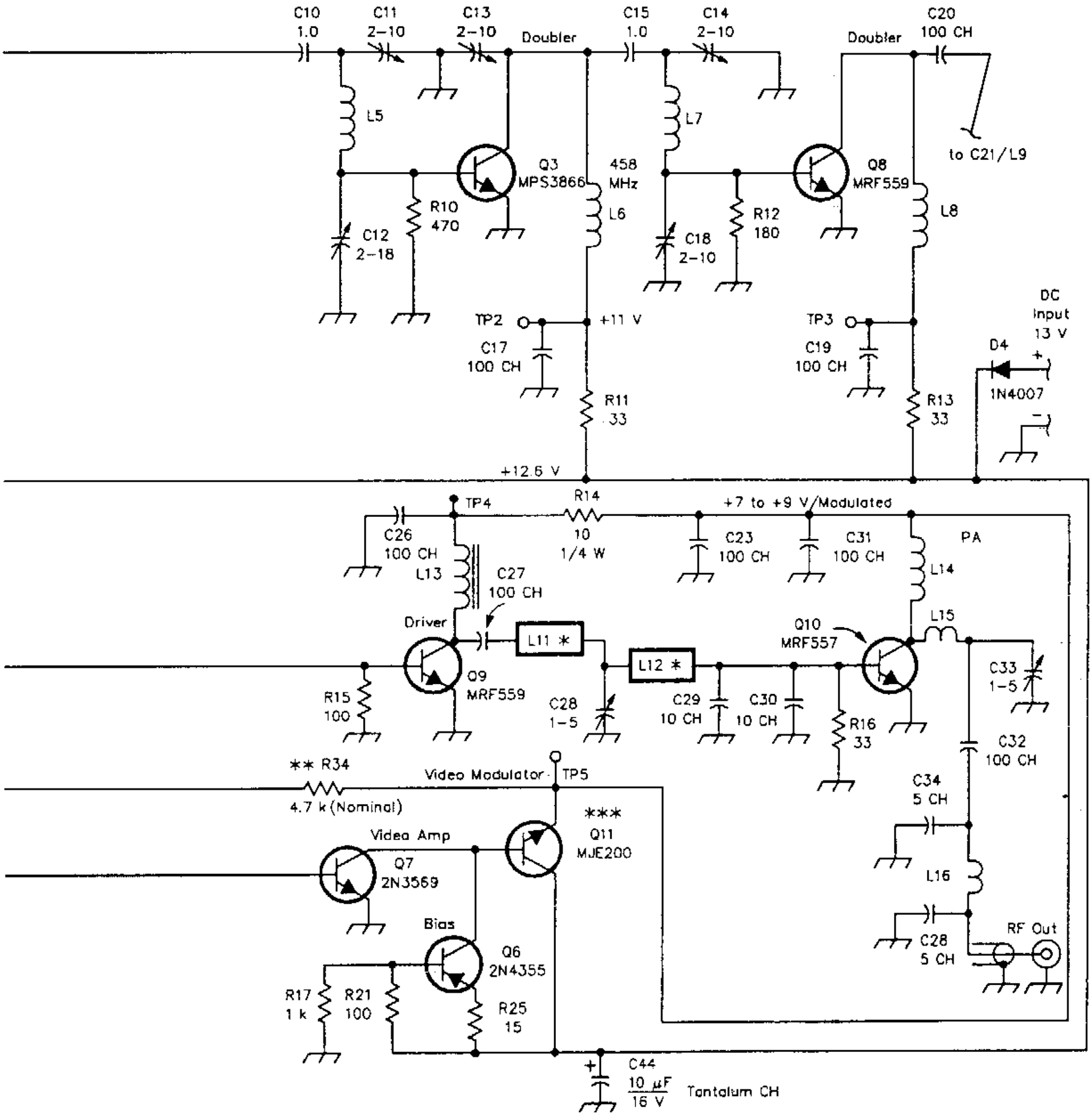
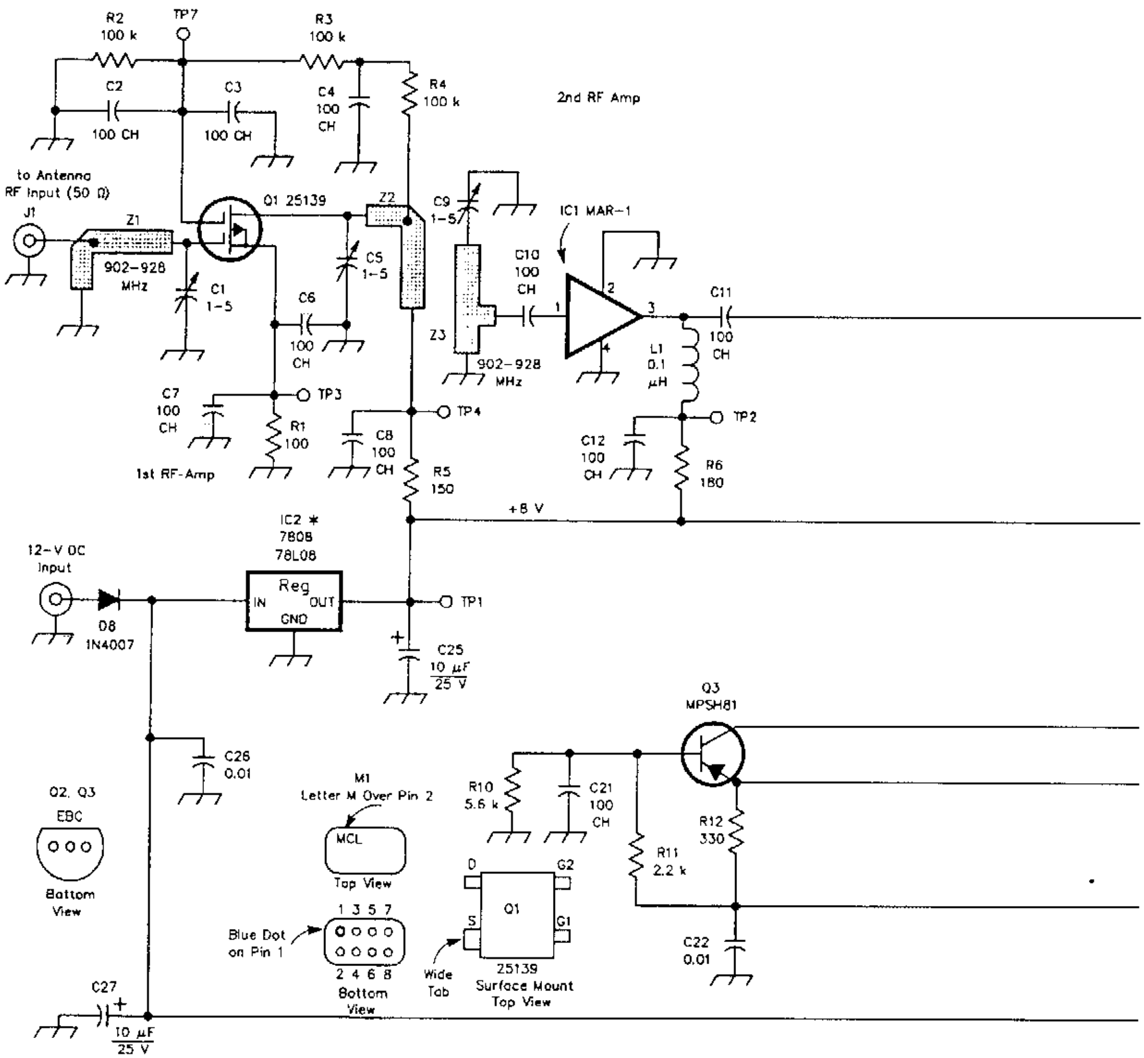


Fig. 5-6

video input. Bandwidth of audio can be restricted to 20 kHz by placing a capacitor with a value of about 0.002 µF across R34. The output is 1.5 to 2 watts PEP into a 50-Ω load. A complete kit of parts including PC board, is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

ATV DOWNCONVERTER FOR 902 TO 928 MHz



QST

This downconverter converts ATV signals in the 902- to 928-MHz range to a 61.25- or 67.25-MHz IF output frequency (CH 3 or CH 4) to enable reception of these signals on a standard VHF TV receiver or monitor. It features a low-noise RF amp feeding a Schottky diode double-balanced mixer, a tunable LO and one IF preamp stage. The RF amplifier is a low-noise dual-gate GASFET that is followed by a second RF stage using an MMIC. Five tuned circuits are used in the RF amplifier. This feeds a packaged Schottky diode mixer assembly for better dynamic range and reduced susceptibility to intermodulation and strong signal areas. The on-board local oscillator (LO) is voltage tuned and if desired can be set up for remote tuning. All necessary circuitry for remote tuning is on board for coax dc and IF feed. This en-

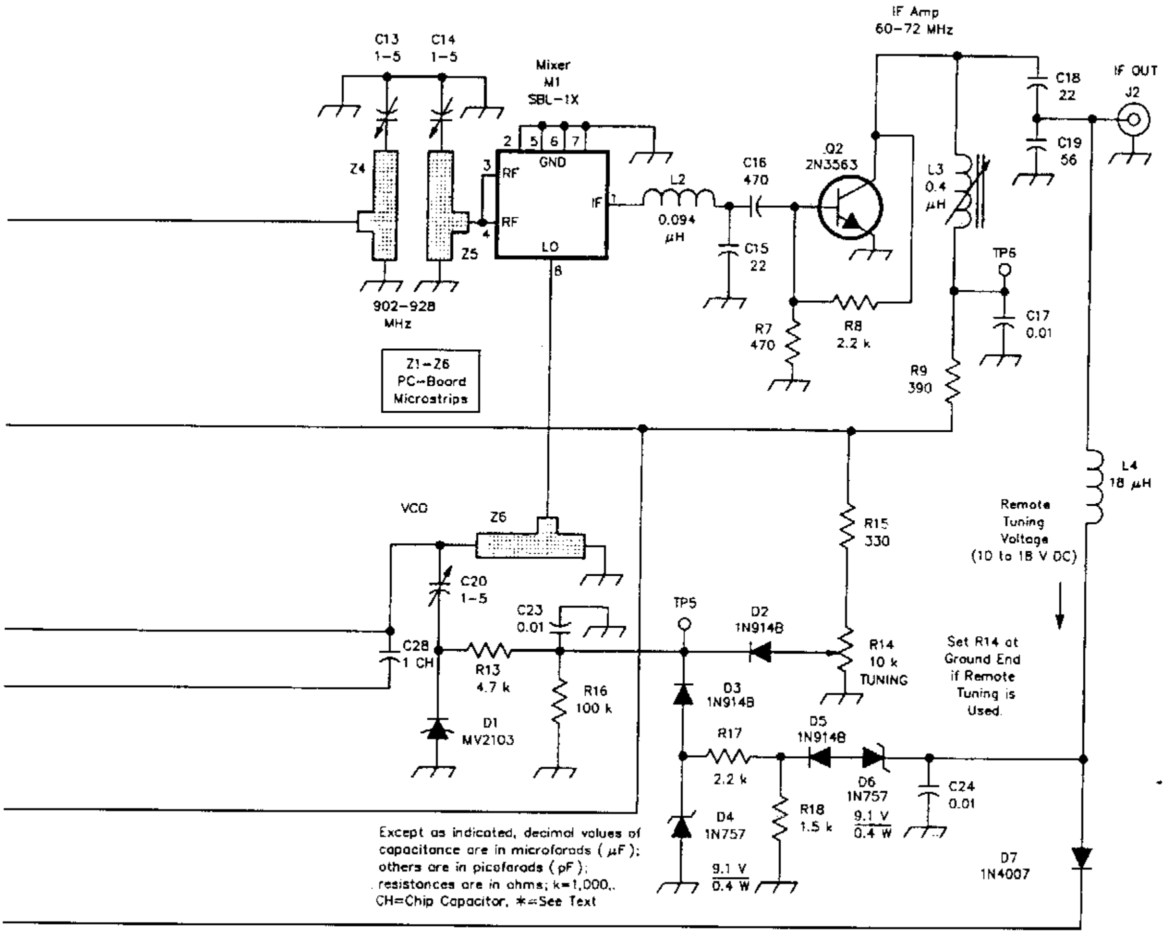
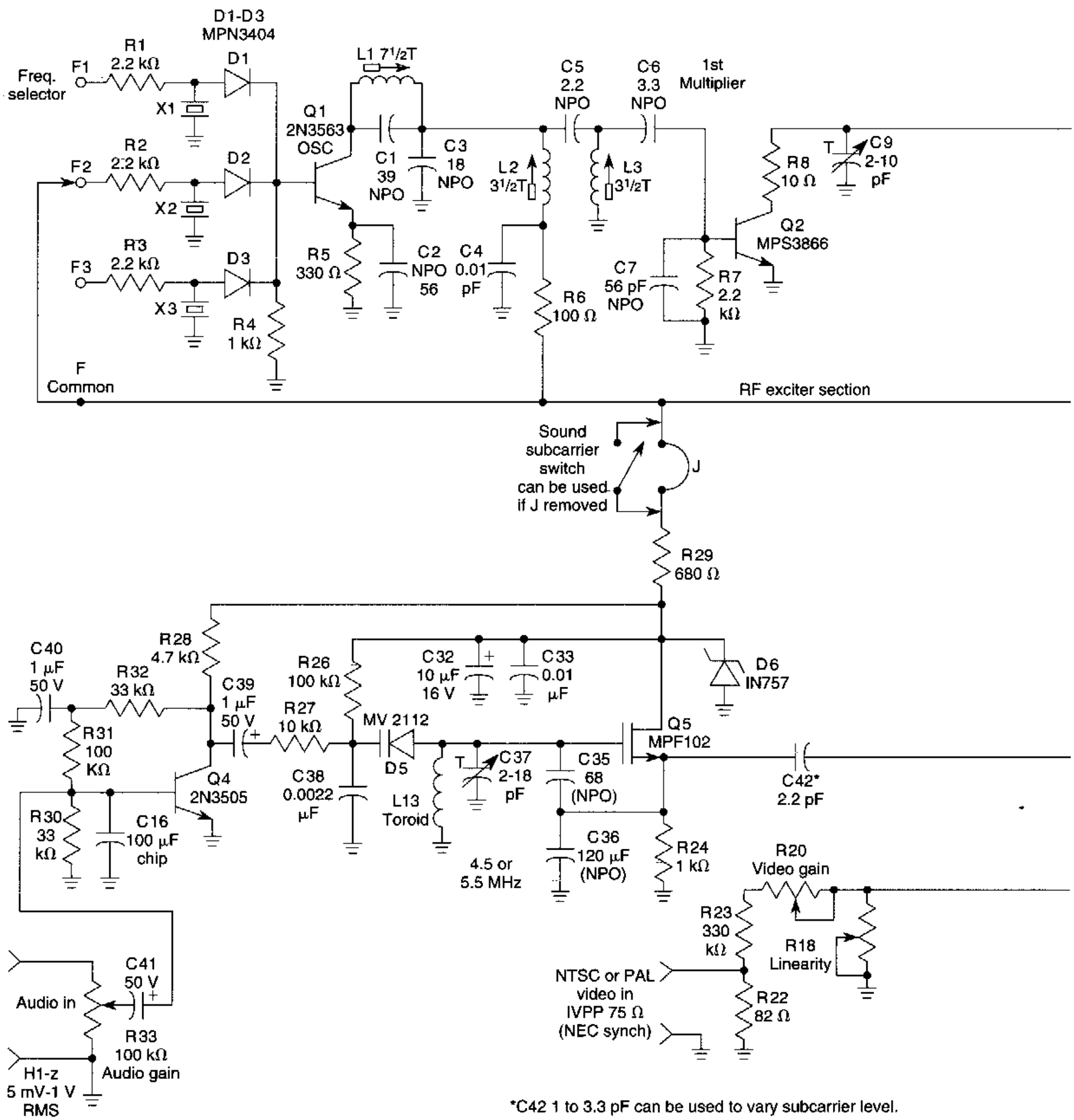


Fig. 5-7

ables the downconverter to be mast mounted to get around feedline losses generally associated with this frequency range. No separate dc feed is necessary because the coax (RG59/U recommended) carries dc power, tuning voltage, and IF signal. A dc block is used at the receiver for the purpose of separating dc voltage supply and the tuning voltage. This allows a cable run of several hundred feet, if needed.

By using this downconverter and transmitter, a physically small 915-MHz ATV station or even a video HT can be constructed because both units are each 2.50 × 4.00 inches × 1.00 high, and can be stacked together. A complete kit of parts, including PC board is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

THREE-CHANNEL 420- TO 450-MHz ATV TRANSMITTER



RUDOLF F. GRAF AND WILLIAM SHEETS

This transmitter is a 2-watt PEP output device for 420- to 450-MHz amateur TV operation. It has three crystal-controlled channels and will accept standard NTSC video input. It also has a 4.5-MHz sound subcarrier capability. Because this transmitter has AM modulation, audio can be transmitted in

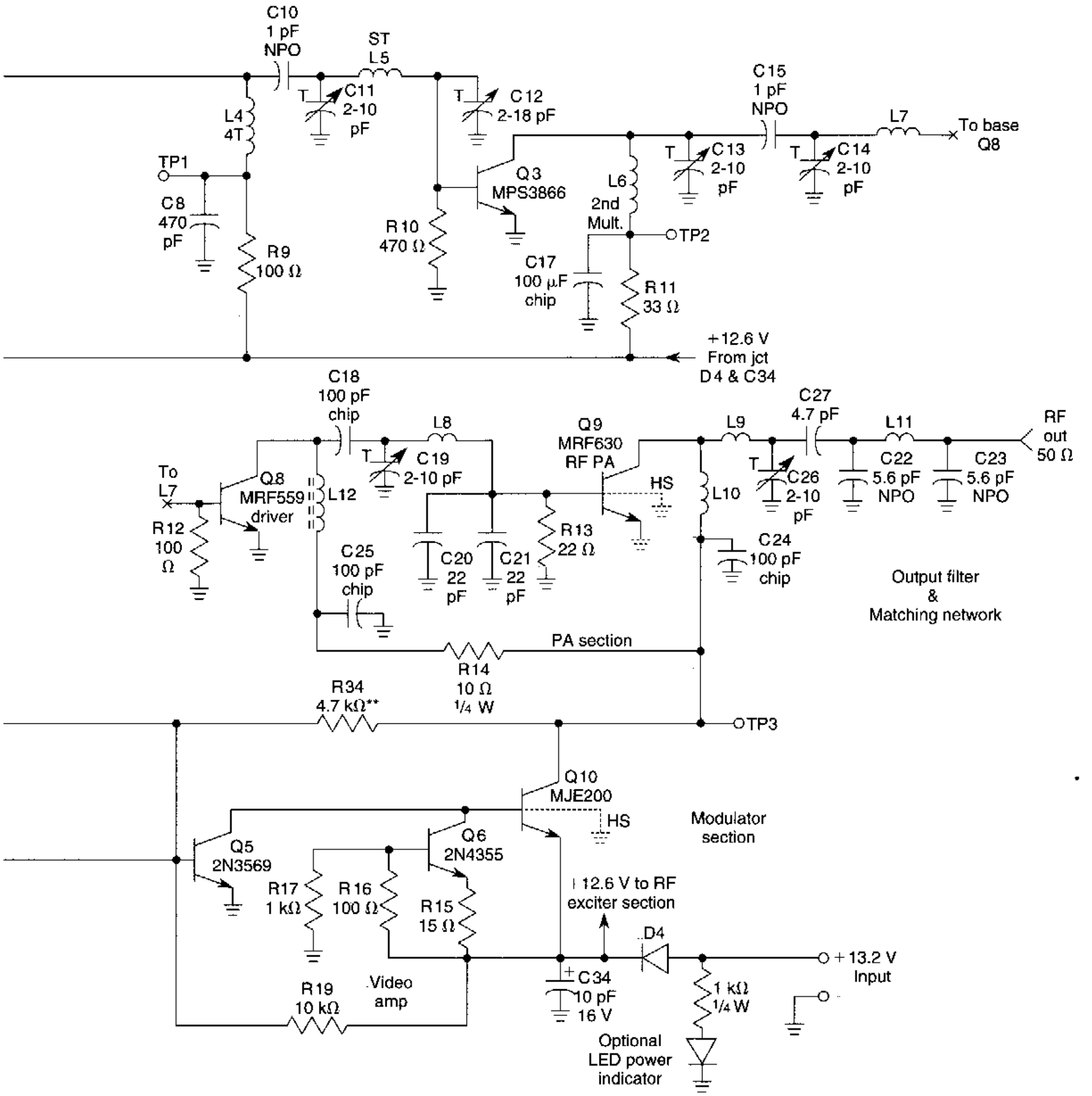
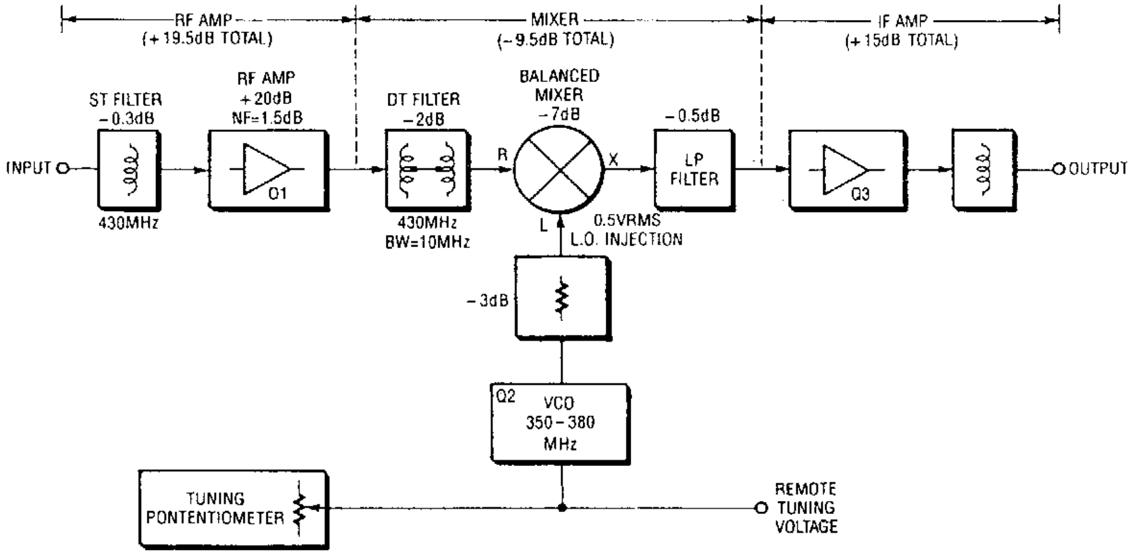


Fig. 5-8

AM form on the RF carrier by applying audio to the video input. Bandwidth of audio can be restricted to 20 kHz by placing a 0.002-µF capacitor across R34. A complete kit of parts, including PC board is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

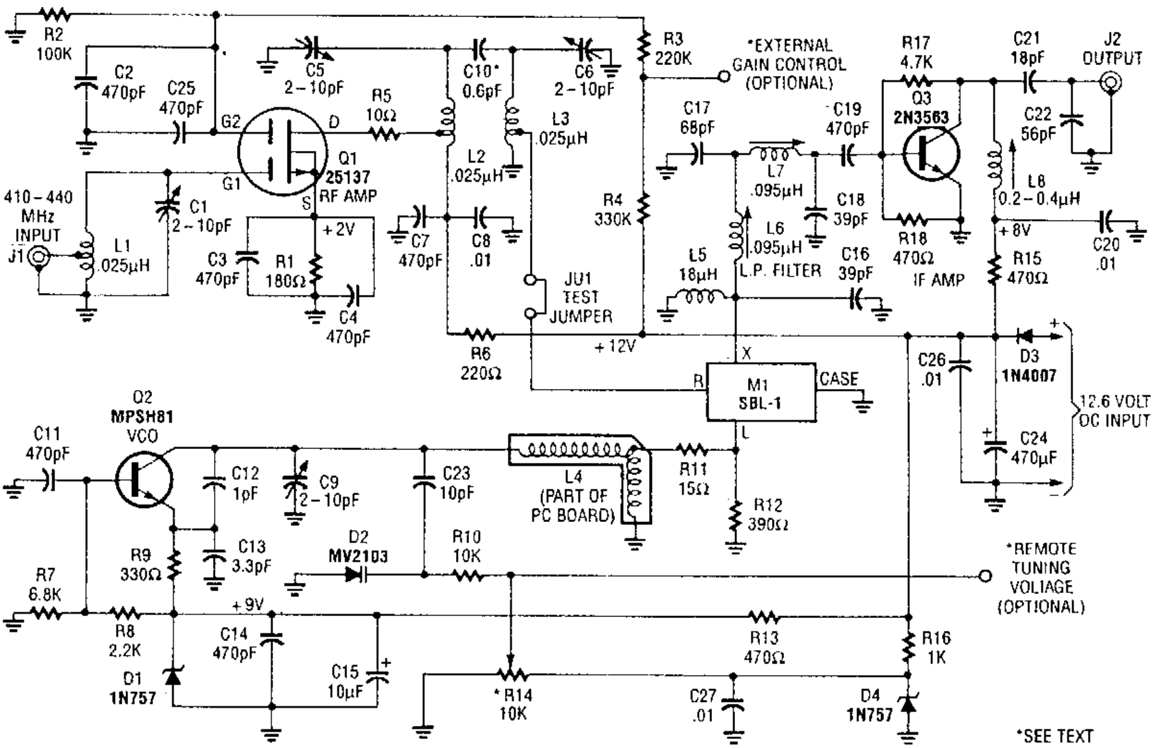
ATV DOWNCONVERTER FOR 420 TO 450 MHz



RADIO-ELECTRONICS *Fig. 5-9*

This ATV downconverter converts the 420- to 450-MHz ATV band, which is several channels below the lower limit of the UHF band, to channel 3 or 4 for viewing on virtually any TV. The downconverter has a low-noise preamplifier stage and a double-balanced passive mixer for good performance and a wide dynamic range. That is necessary with today's crowded UHF bands. The converter draws about 27 milliamperes from a 13.2-volt dc source, so it can be used in portable and mobile applications. An extra IF stage gives an overall gain of about 25 dB. A block diagram of the downconverter is also shown. A complete kit of parts, including PC board, is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804-0053A.

ATV DOWNCONVERTER FOR 420 TO 450 MHz (Cont.)

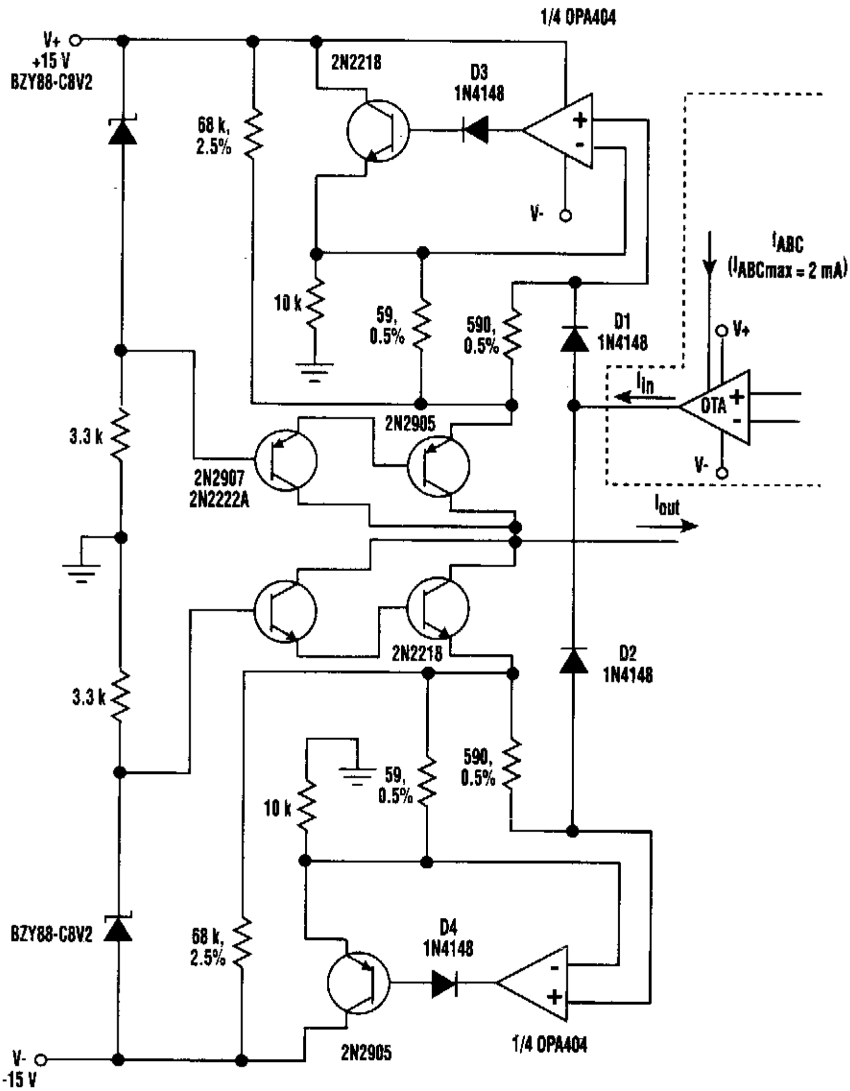


6

Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

OPERATIONAL TRANSCONDUCTANCE AMPLIFIER WITH BOOSTER

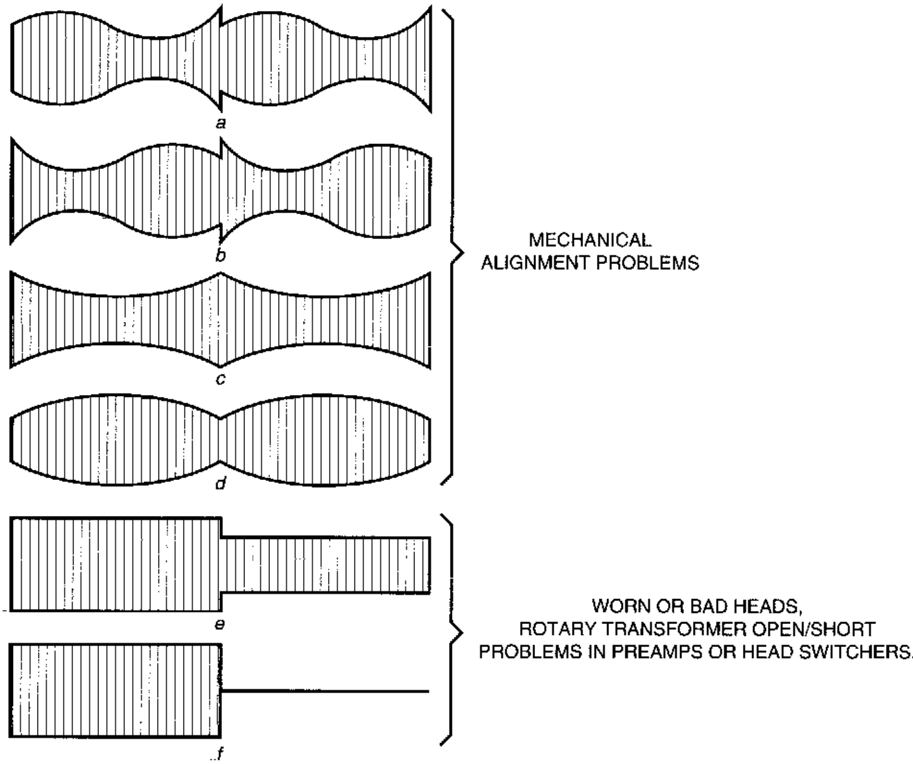


ELECTRONIC DESIGN

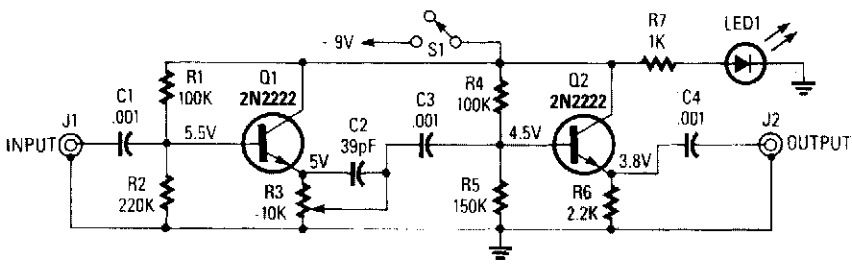
Fig. 6-1

Implementing a bidirectional precision current amplifier in an operational transconductance amplifier (OTA) can boost the OTA's output current. To accomplish this task, two diodes and a complementary stage are added to this otherwise simple design.

VCR HEAD AMPLIFIER TESTER



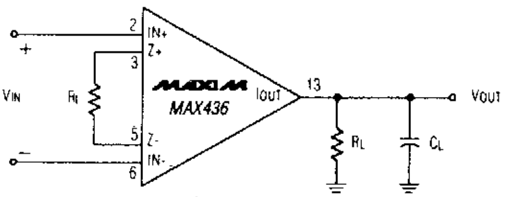
IMPROPER WAVEFORMS. Waveforms *a-d* are caused by mechanical misalignment of the tape guides. The waveforms in *e* and *f* indicate proper alignment, but show that there's a problem with either the video heads, pre-amps, or head switcher.



R-E EXPERIMENTERS HANDBOOK ***Fig. 6-2***

This amplifier enables you to use a signal from a working VCR to test the head amplifiers of a suspected defective VCR. The circuit is basically a video amplifier.

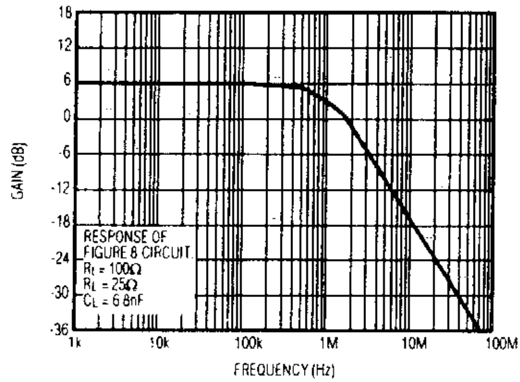
LOWPASS AMPLIFIER



POLE FREQUENCY = $F_P = \frac{1}{2\pi R_L C_L}$

PASSBAND GAIN = $K\left(\frac{R_L}{R_t}\right)$

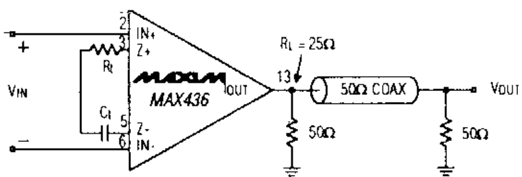
Lowpass Amplifier



Lowpass Amplifier Gain vs. Frequency

MAXIM ***Fig. 6-3***

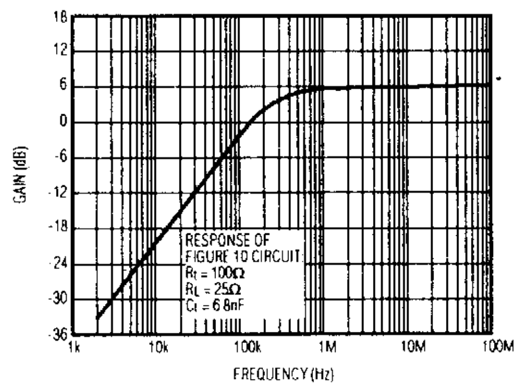
HIGHPASS AMPLIFIER



CORNER FREQUENCY = $F_C = \frac{1}{(2\pi)(R_t)(C_t)}$

PASSBAND GAIN = $K\left(\frac{R_L}{R_t}\right)$

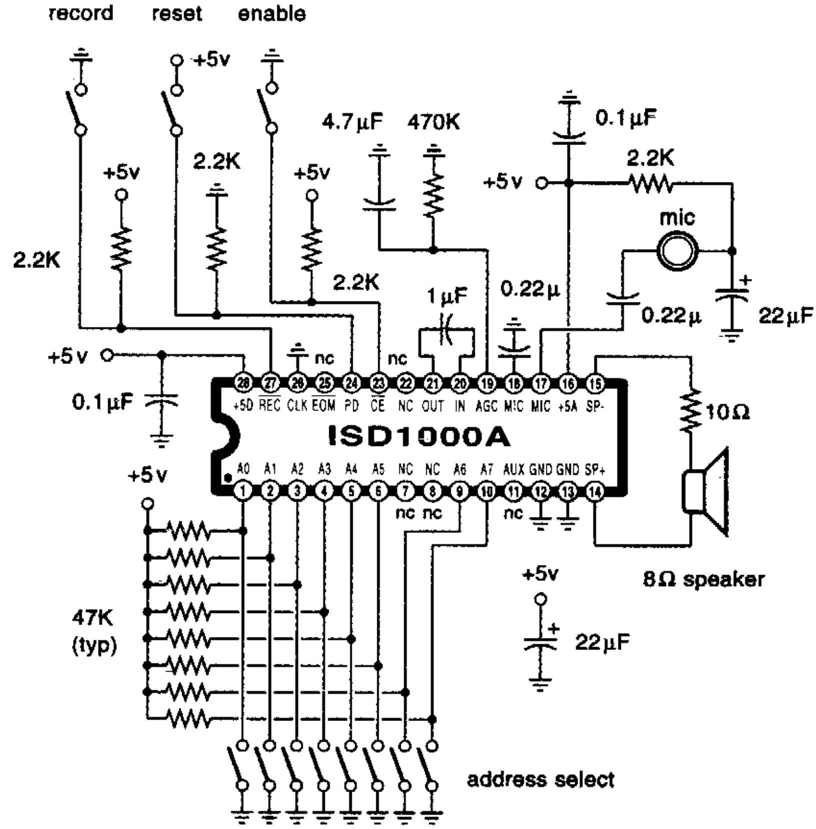
Highpass Amplifier



Highpass Amplifier Gain vs. Frequency

MAXIM ***Fig. 6-4***

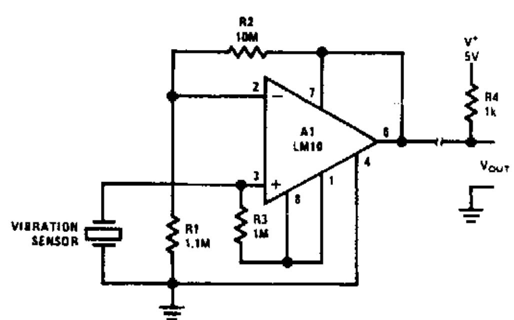
ISD 1000A RECORD/PLAYBACK CIRCUIT



ELECTRONICS NOW *Fig. 6-5*

This circuit uses the Information Storage Devices ISD1000A chip (Radio Shack P/N 276-1325).

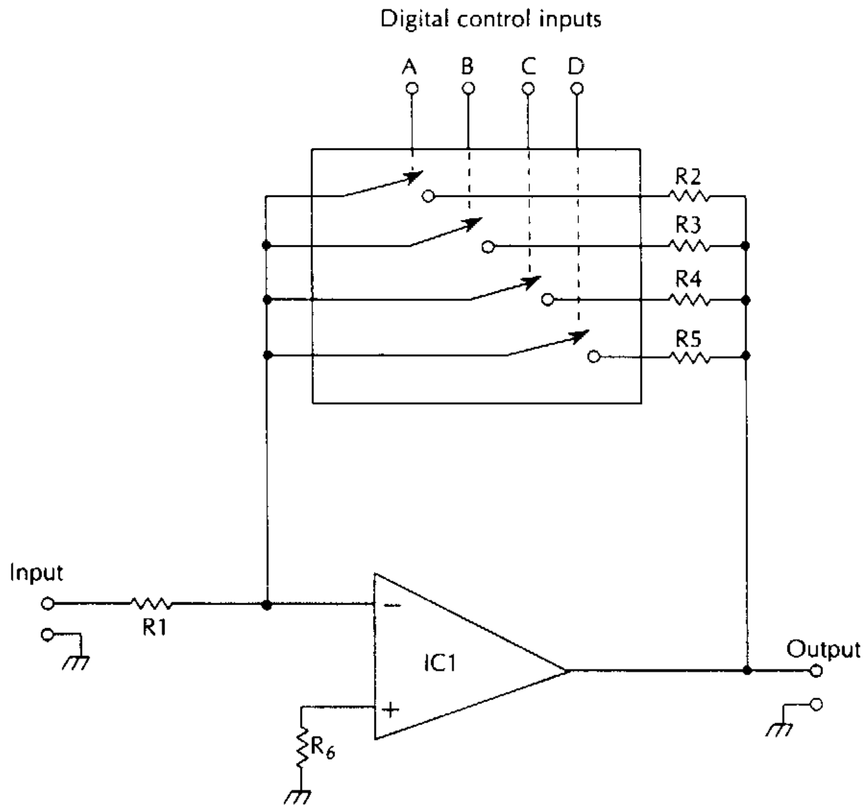
REMOTE AMPLIFIER



Useful for transducers and such where a single two-wire pair are the only leads available.

NATIONAL SEMICONDUCTOR *Fig. 6-6*

PROGRAMMABLE GAIN AMPLIFIER

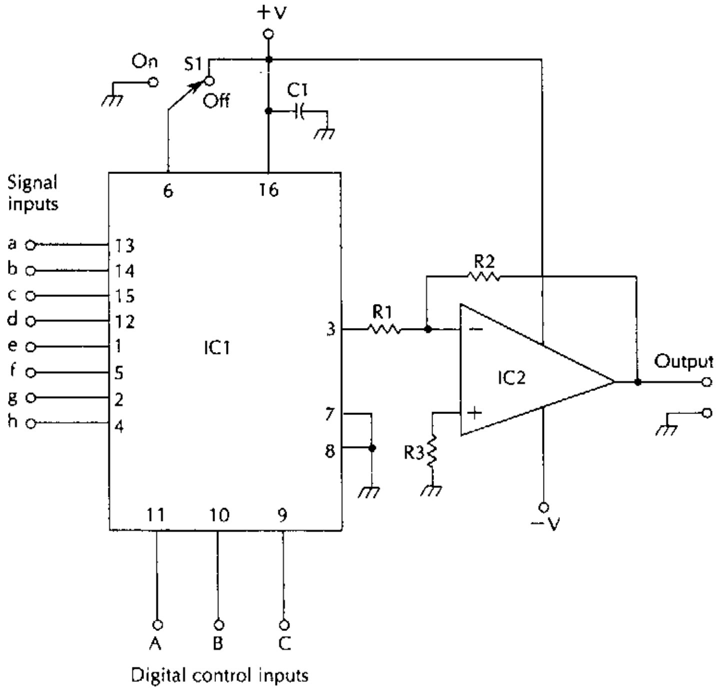


AMPLIFIERS, WAVEFORM GENERATORS & OTHER LOW-COST IC PROJECTS ***Fig. 6-7***

The gain of this amplifier is $-R_f/R_1$ where R_f = effective value of resistance selected by the digital R1 inputs.

IC1	op amp
IC2	CD4066 quad bilateral switch
R1	1-kΩ, ¼-W 5% resistor
R2	10-kΩ, ¼-W 5% resistor
R3	4.7-kΩ, ¼-W 5% resistor
R4	2.2-kΩ, ¼-W 5% resistor
R5	1-kΩ, ¼-W 5% resistor
R6	2.2-kΩ, ¼-W 5% resistor

PROGRAMMABLE INPUT AMPLIFIER

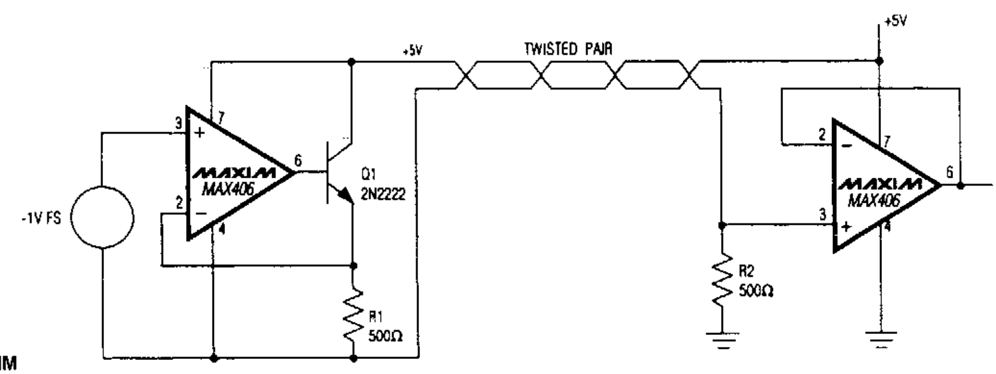


AMPLIFIERS, WAVEFORM GENERATORS & OTHER LOW-COST IC PROJECTS *Fig. 6-8*

This amplifier has eight inputs selectable digitally.

IC1	CD4051 SP8T bilateral switch	R1	10-kΩ, ¼-W 5% resistor
IC2	op amp to suit application	R2	22-kΩ, ¼-W 5% resistor
C1	0.1-µF capacitor	R3	18-kΩ, ¼-W 5% resistor
S1	SPST switch		

REMOTELY POWERED SENSOR AMPLIFIER



MAXIM *Fig. 6-9*

For remote sensor applications, this circuit enables use of a single twisted pair.

TUNED AMPLIFIER

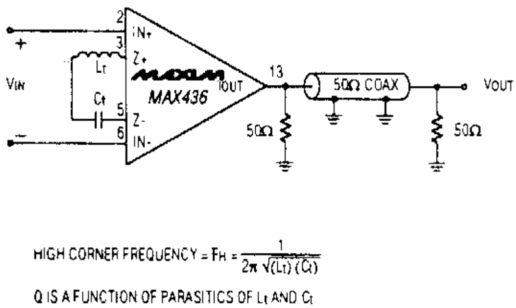


Figure A Tuned Amplifier

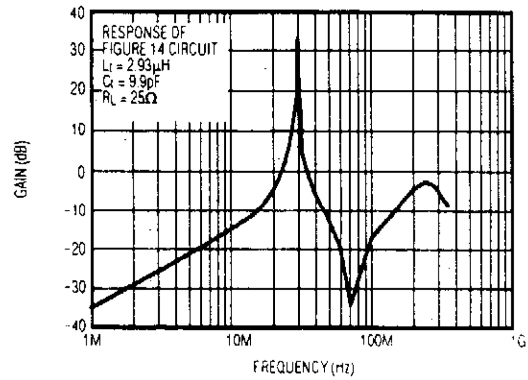


Figure B Tuned Amplifier Gain vs. Frequency

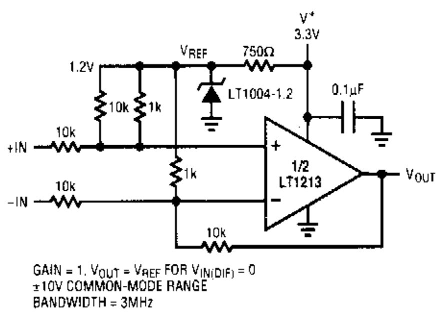
MAXIM ***Fig. 6-10***

This circuit is a tuned amplifier circuit, tuned to the resonant frequency of the LC transconductance network:

$$F_c = \frac{1}{2\pi \sqrt{L_t C_t}}$$

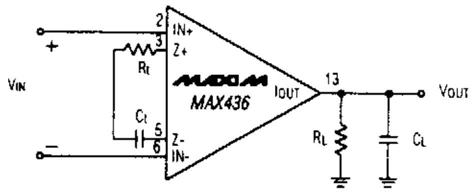
The impedance of the transconductance network is a minimum at the resonant frequency, providing maximum amplifier gain at that frequency. The Q of the amplifier is a function of the parasitic components associated with the LC network. The graph is the frequency response of the circuit, with L_t = 2.93 μH and C_t = 9.9 pF.

DIFFERENCE AMPLIFIER WITH WIDE INPUT COMMON-MODE RANGE



LINEAR TECHNOLOGY ***Fig. 6-11***

BANDPASS AMPLIFIER



LOW CORNER FREQUENCY = $F_L = \frac{1}{(2\pi)(R_t)(C_t)}$

POLE FREQUENCY = $F_P = \frac{1}{(2\pi)(R_L)(C_L)}$

PASSBAND GAIN = $K\left(\frac{R_L}{R_t}\right)$

Figure A Bandpass Amplifier

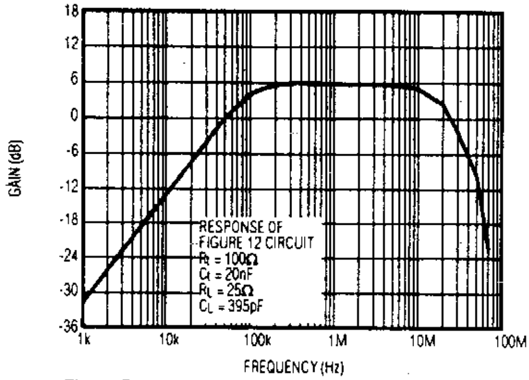
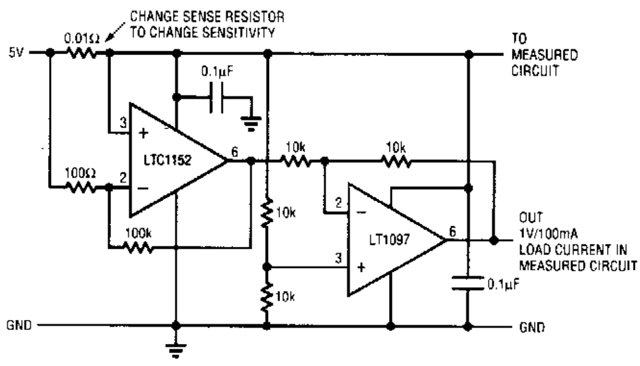


Figure B

MAXIM ***Fig. 6-12***

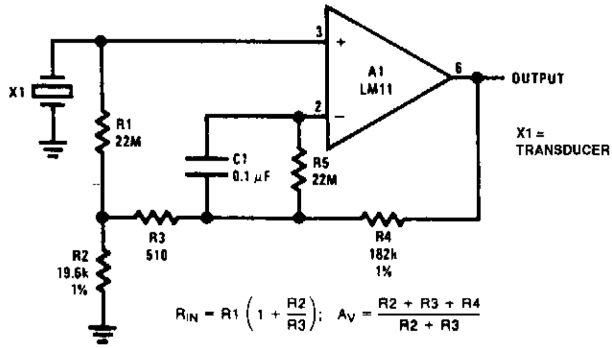
The circuit A is a bandpass amplifier, with the low corner frequency set by the impedance of the transconductance network. The high corner frequency is set by the impedance of the RC network at the amplifier output. The passband gain is (k) x (R_l/R_t). Figure B is a plot of the circuit in Figure A, with R_t = 100 Ω, C_t = 20 nF, R_l = 25 Ω, and C_l = 395 pF.

HIGH-SIDE CURRENT-SENSING AMPLIFIER



LINEAR TECHNOLOGY ***Fig. 6-13***

HIGH INPUT IMPEDANCE ac AMPLIFIER

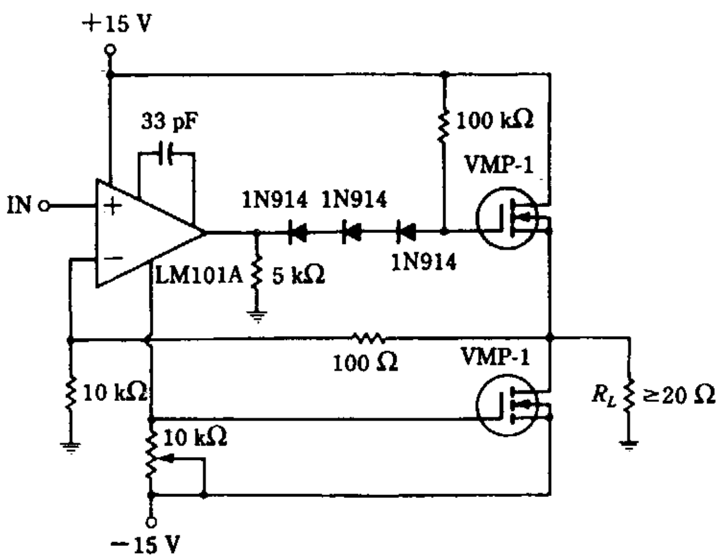


NATIONAL SEMICONDUCTOR — *Fig. 6-14*

This figure shows an op amp used as an ac amplifier. It is unusual in that dc bootstrapping is used to obtain high input resistance without requiring high-value resistors. In theory, this increases the output offset because the op amp offset voltage is multiplied by the resistance boost.

But when conventional resistor values are used, it is practical to include R5 to eliminate bias-current error. This gives less output offset than if a single, large resistor were used. C1 is included to reduce noise.

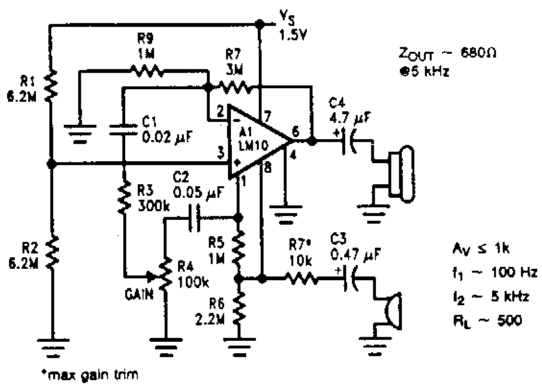
MOSFET PUSH-PULL AMPLIFIER



MCGRAW-HILL — *Fig. 6-15*

This amplifier can be used for audio or as a driver for inverter service.

LOW-VOLTAGE MICROPHONE PREAMP



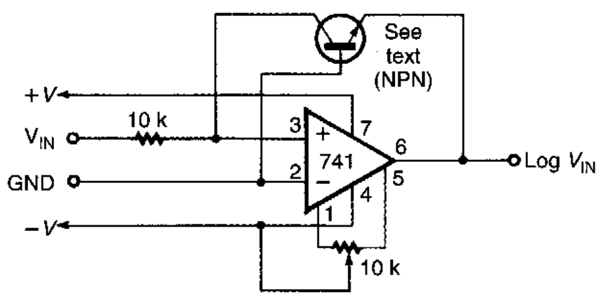
A microphone amplifier is shown. The reference, with a 500-kHz unity-gain bandwidth, is used as a preamplifier with a gain of 100. Its output is fed through a gain-control potentiometer to the op amp, which is connected for a gain of 10. The combination gives a 60-dB gain with a 10-kHz bandwidth, unloaded, and 5 kHz loaded at 500 Ω. Input impedance is 10 kΩ.

Potentially, using the reference as a preamplifier in this fashion can cause excess noise. However, because the reference voltage is low, the noise contribution, which adds root-mean-square, is likewise low. The input noise voltage in this connection is 440-500 nV Hz, about equal to that of the op amp.

NATIONAL SEMICONDUCTOR

Fig. 6-16

BASIC LOGARITHMIC AMPLIFIER USING OP AMP

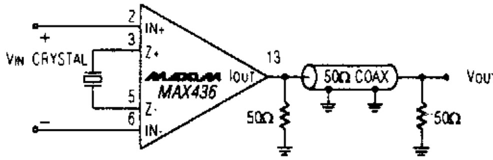


ELECTRONICS NOW!

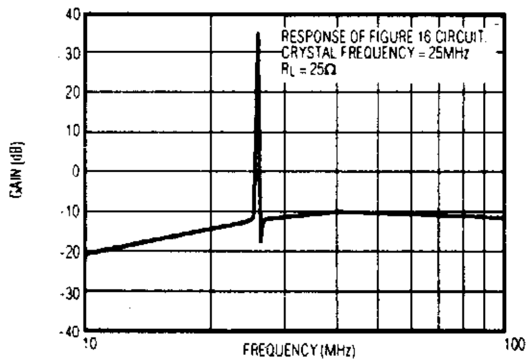
Fig. 6-17

This logarithmic amplifier uses a single op amp. The current in the feedback loop of the op amp is equal to the current flow at the input of the op amp.

CRYSTAL TUNED AMPLIFIER



Crystal Tuned Amplifier



Crystal Tuned Amplifier Gain vs. Frequency

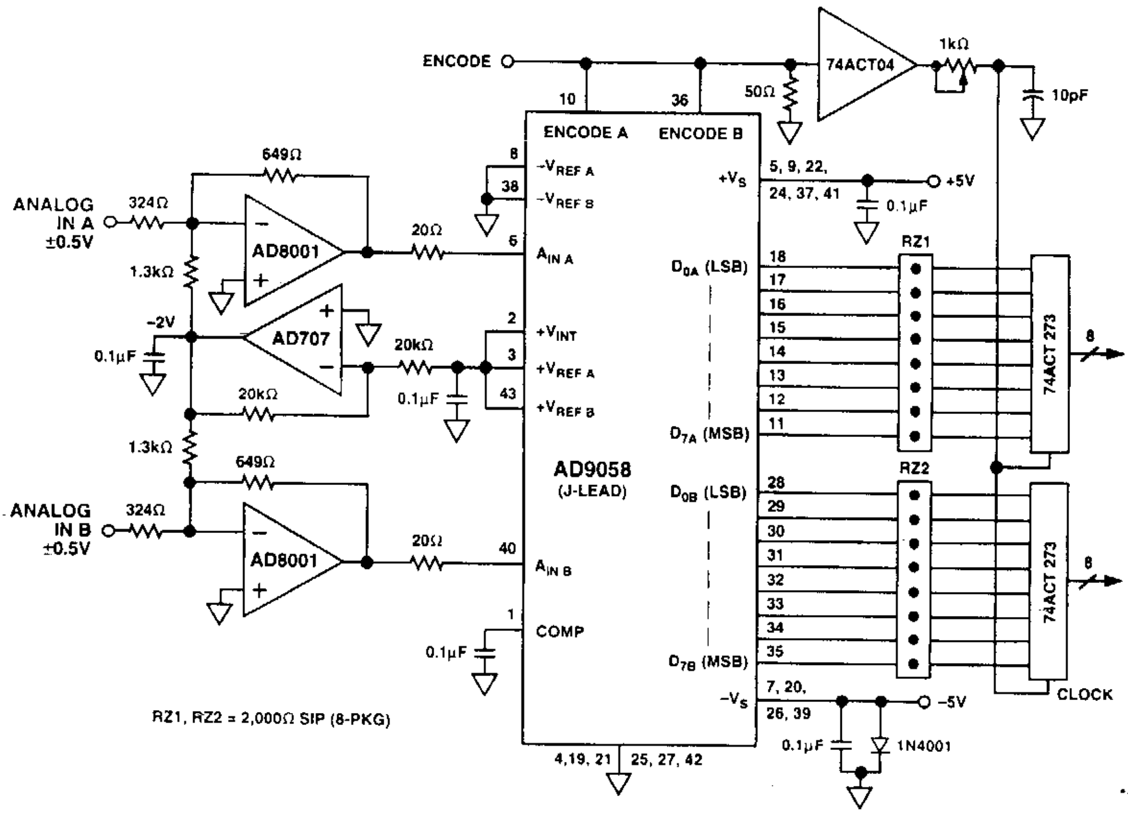
MAXIM ***Fig. 6-18***

7

A/D Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

HIGH-SPEED A/D CONVERTER SYSTEM

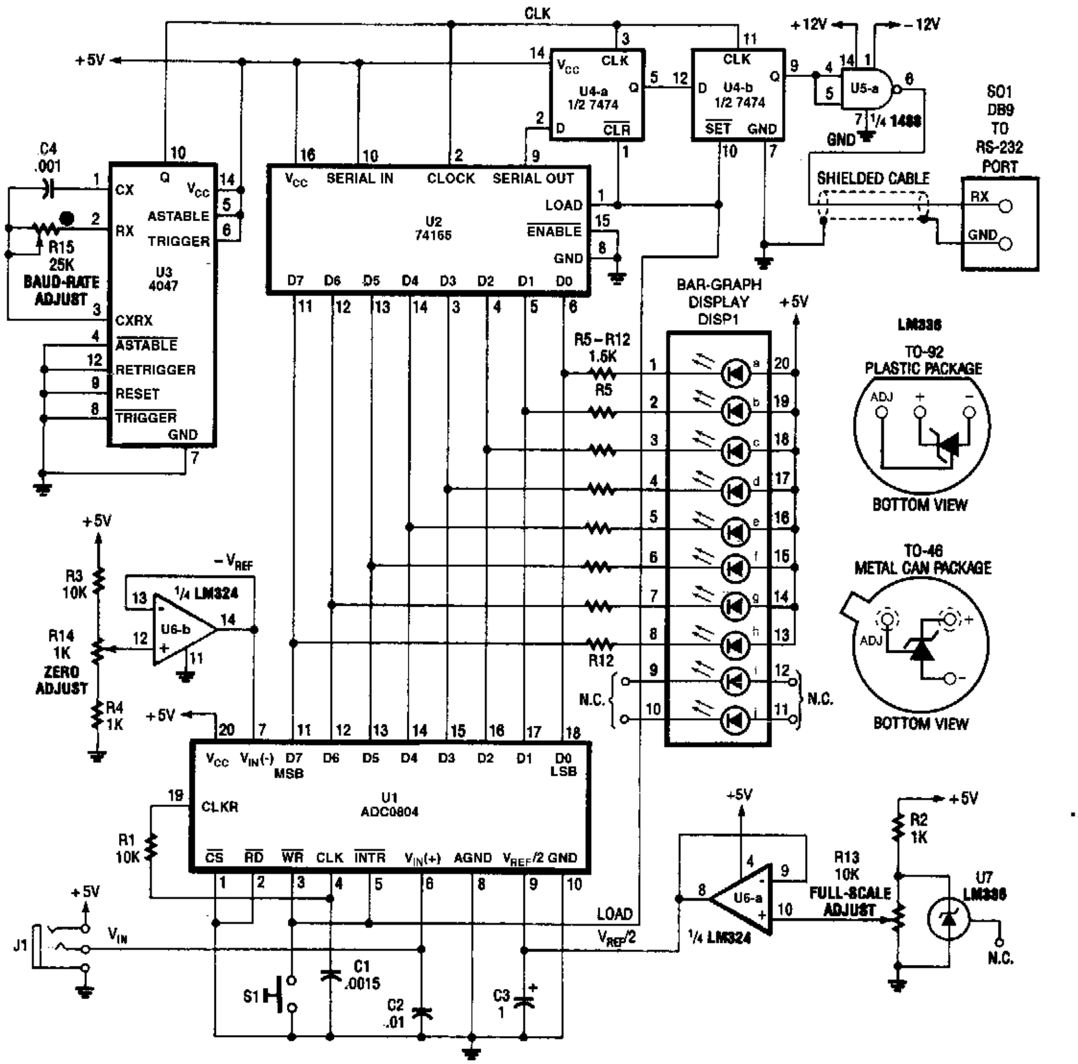


ANALOG DEVICES

Fig. 7-1

The AD8001 is well suited for driving high-speed analog-to-digital converters, such as the AD9058. The AD9058 is a dual 8-bit 50 Msps ADC. In the circuit shown, there are two AD8001s driving the inputs of the AD905f8 which are configured for 0- to +2-V ranges. Bipolar input signals are buffered, amplified (–2×), and offset (by +1.0 V) into the proper input range of the ADC. Using the AD9058's internal +2-V reference connected to both ADCs (as shown) reduces the number of external components required to create a complete data acquisition system. The 20-Ω resistors in series with ADC input are used to help the AD8001 drive the 10-pF ADC input capacitance. The two AD8001s only add 100 mW to the power consumption while not limiting the performance of the circuit.

A/D CONVERTER FOR PCs



POPULAR ELECTRONICS *Fig. 7-2*

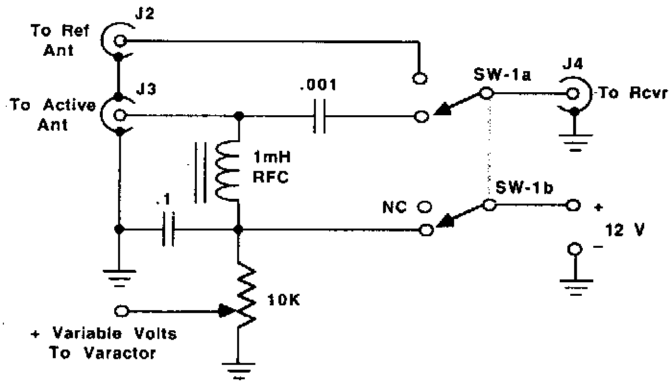
An ADC0804 A/D converter converts analog data to digital. This is fed to a 74165 8-bit shift register and converted to serial data. U3 provides a baud-rate clock. U4A and U4B are used to generate start and stop bits needed at beginning and end of each data word.

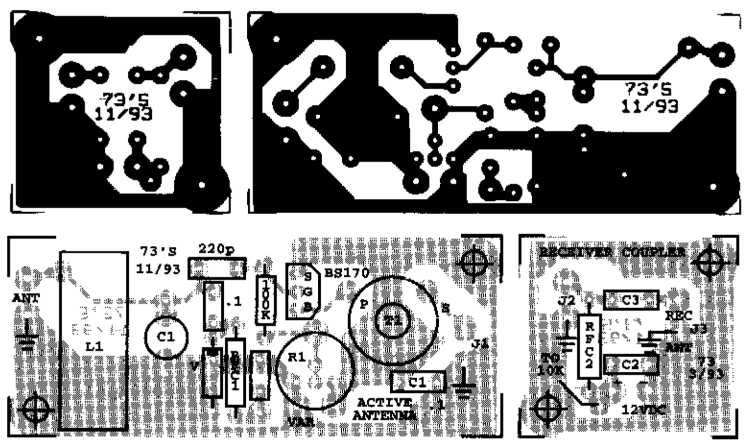
8

Antenna Circuits

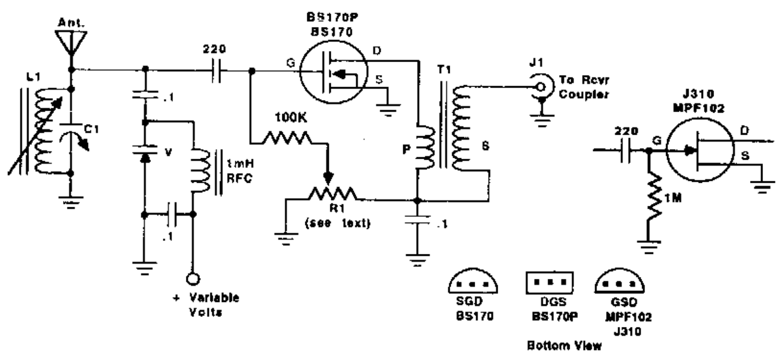
The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

REMOTE TUNED ACTIVE HF ANTENNA





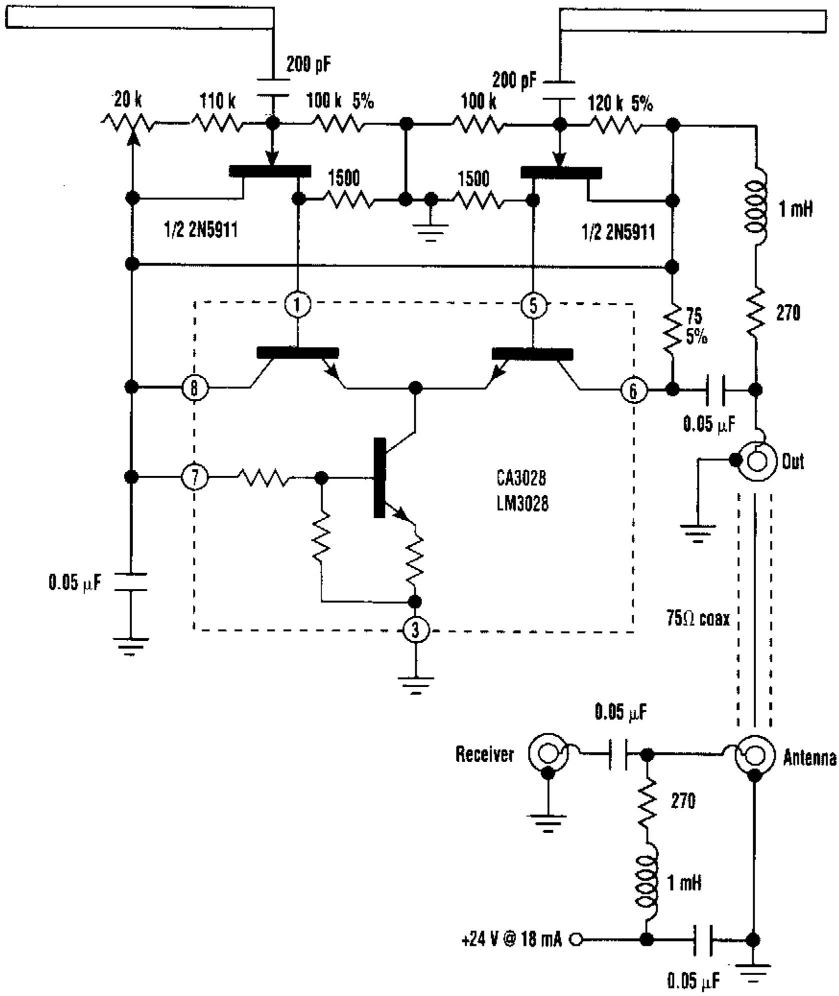
PC board pattern and parts placement diagram.



73 AMATEUR RADIO TODAY ***Fig. 8-1***

An MV1662/S varactor diode tunes this active antenna/preamplifier. R1 varies gate bias on the BS170 FET. T1 is a 3:1 toroidal winding suitable for the frequencies of interest.

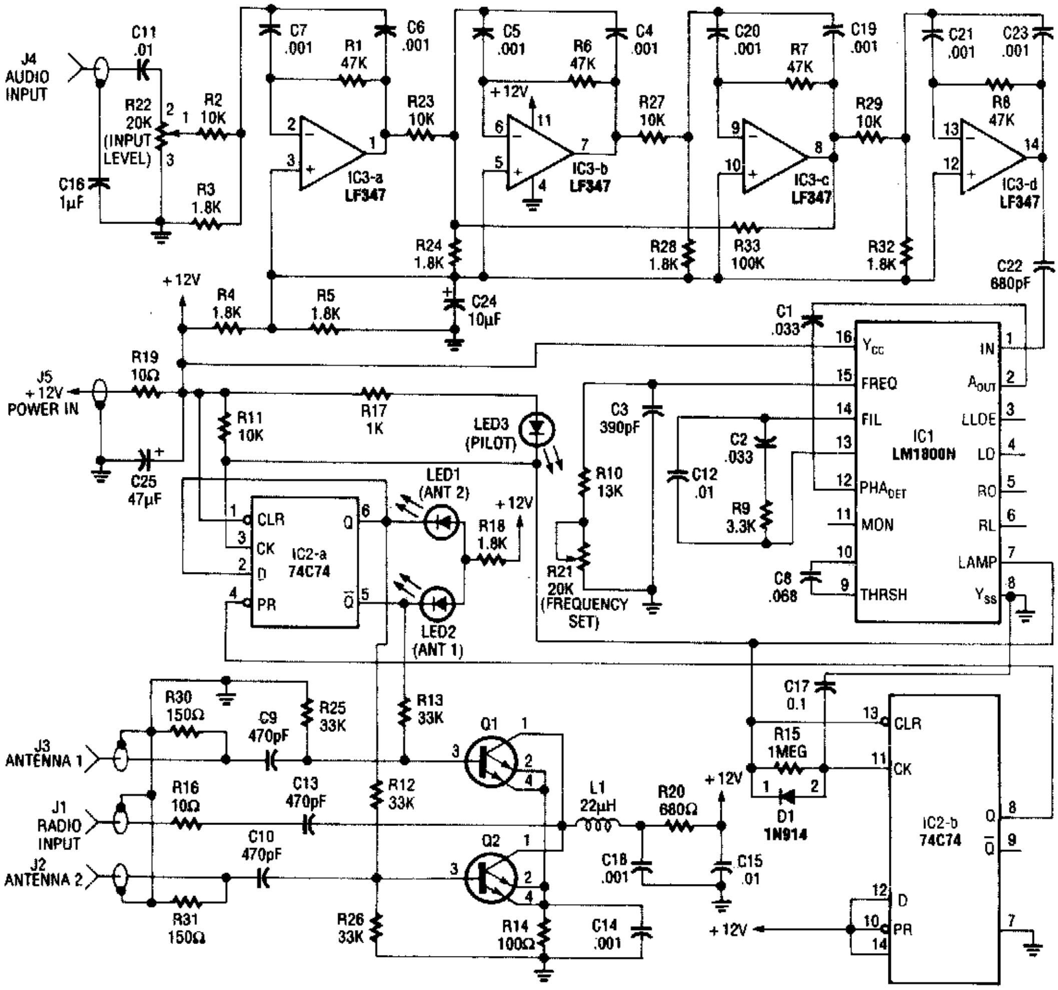
MINIATURE BROADBAND ANTENNA (3 TO 30 MHz)



ELECTRONIC DESIGN *Fig. 8-2*

A short dipole antenna and impedance converter combined together can be rotated to null out an interfering signal. The converter supplies a tremendous current gain so that the voltage appearing at the dipole's output eventually drives a 75-Ω load.

FM AUTO RADIO DIVERSITY ANTENNA



ELECTRONICS NOW *Fig. 8-3*

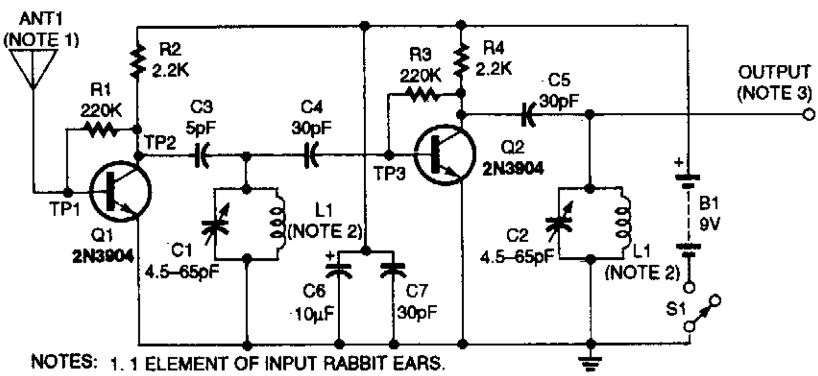
A second antenna, installed on your vehicle as far away from the original equipment antenna as practical, provides the second FM signal. The figure is a simplified block diagram of the diversity system.

The cables from both antennas are connected to the electronic antenna switch. The 19-kHz pilot signal from the receiver's audio output is passed through a high-gain bandpass active filter, which attenuates audio programming that is much stronger than the pilot signal. After amplification, the pilot subcarrier becomes the reference frequency for a phase-locked loop (PLL) circuit. The output of the PLL locks to the 19-kHz pilot signal and functions as a subcarrier detector. When the reference frequency becomes noisy, the PLL will lose "lock" and trigger the flip-flop, whose output switches the state of the electronic antenna switch. This action switches the alternate antenna into the system while disabling the original antenna.

If that second antenna is positioned for better reception, the received signal will clear, and the PLL will again lock to the subcarrier and hold the switch in that state until the pilot signal drops out again. If the second antenna does not restore the pilot signal reception after a 0.1-second delay, the primary antenna is switched back on.

When the radio is receiving AM, the absence of the 19-kHz subcarrier will also reactivate the primary antenna that is tuned to the receiver for the best AM reception.

TUNABLE FM ANTENNA BOOSTER

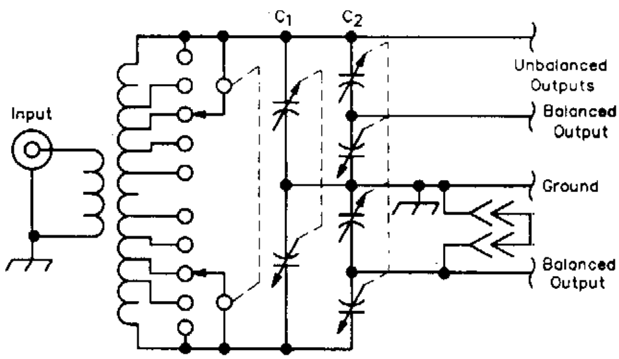


NOTES: 1. 1 ELEMENT OF INPUT RABBIT EARS.
2. SEE TEXT FOR WINDING INSTRUCTIONS
3. TO ANTENNA OF FM RECEIVER
4. VOLTAGES AT TEST POINTS: TP1=0.68V, TP2=3.85V, TP3=0.68V.

ELECTRONICS NOW — *Fig. 8-4*

This two-transistor amplifier circuit with tunable tank circuits boosts the distant FM signals. Coils L1 and L2 are 1½ turns #20 AWG bail tinned wire wound around a ⅜" diameter mandrel.

MATCHBOX ANTENNA TUNER

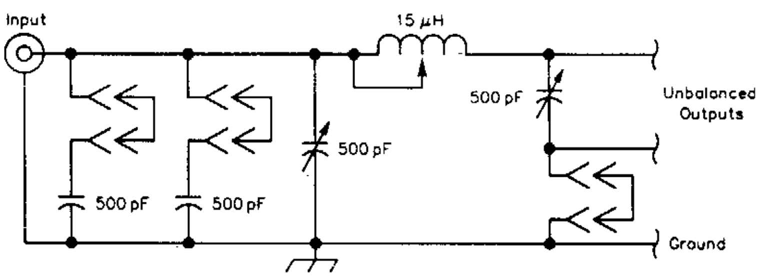


QST

Fig. 8-5

C1 is a split stator capacitor and C2 is a dual differential capacitor. The top unbalanced output connection is used for high-impedance unbalanced loads, and the other is used for low-impedance unbalanced loads. In the latter case, the unused balanced load connection is grounded.

ANTENNA TUNER

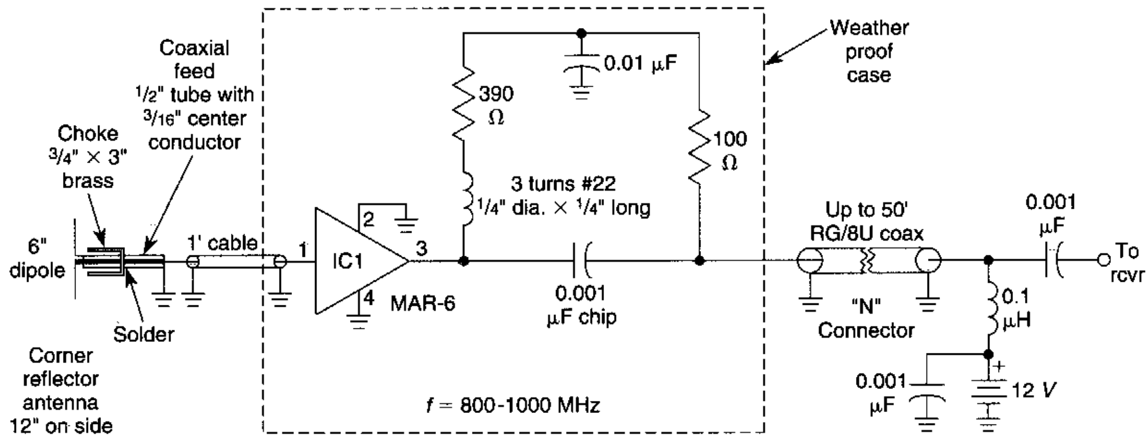


QST

Fig. 8-6

This is a circuit diagram of the Collins Model 180S-1 antenna tuner. Three unbalanced configurations are available, two of which form an L-network and the other is a π-network. The tuning range is impressive.

ACTIVE ANTENNA FOR UHF SCANNERS



WILLIAM SHEETS *Fig. 8-7*

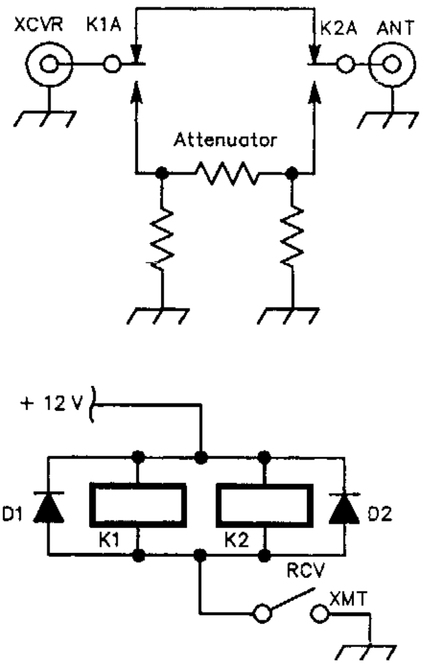
This active antenna is a ½-wave dipole mounted in a 12" × 12" × 12" 90° corner reflector. A built-in active preamp IC1, fed dc through the RF coaxial line, provides 15 dB gain at 900 MHz to offset cable losses. This provides superior reception for scanners covering the 800- to 1000-MHz range.

9

Attenuator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SWITCHABLE POWER ATTENUATOR



QST

Fig. 9-1

Schematic diagram of a switchable power attenuator that can be used to reduce the power output of transmitters that don't have ALC lines.

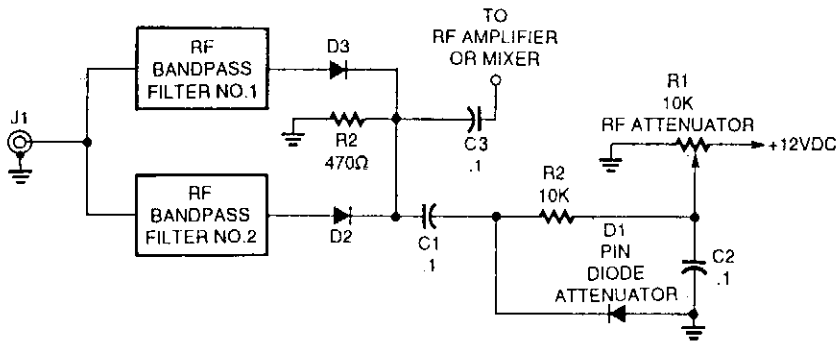
Values for 10 and 20 dB:

10 dB: $R_1 = R_3 = 91\ \Omega$
$R_2 = 75\ \Omega$ nearest standard values

20 dB: $R_1 = R_3 = 62\ \Omega$
$R_2 = 240\ \Omega$ nearest standard values

Note: R1 must handle the largest share of the input power, and R2 somewhat less. This depends on attenuation selected.

VARIABLE VOLTAGE ATTENUATOR



POPULAR ELECTRONICS ***Fig. 9-2***

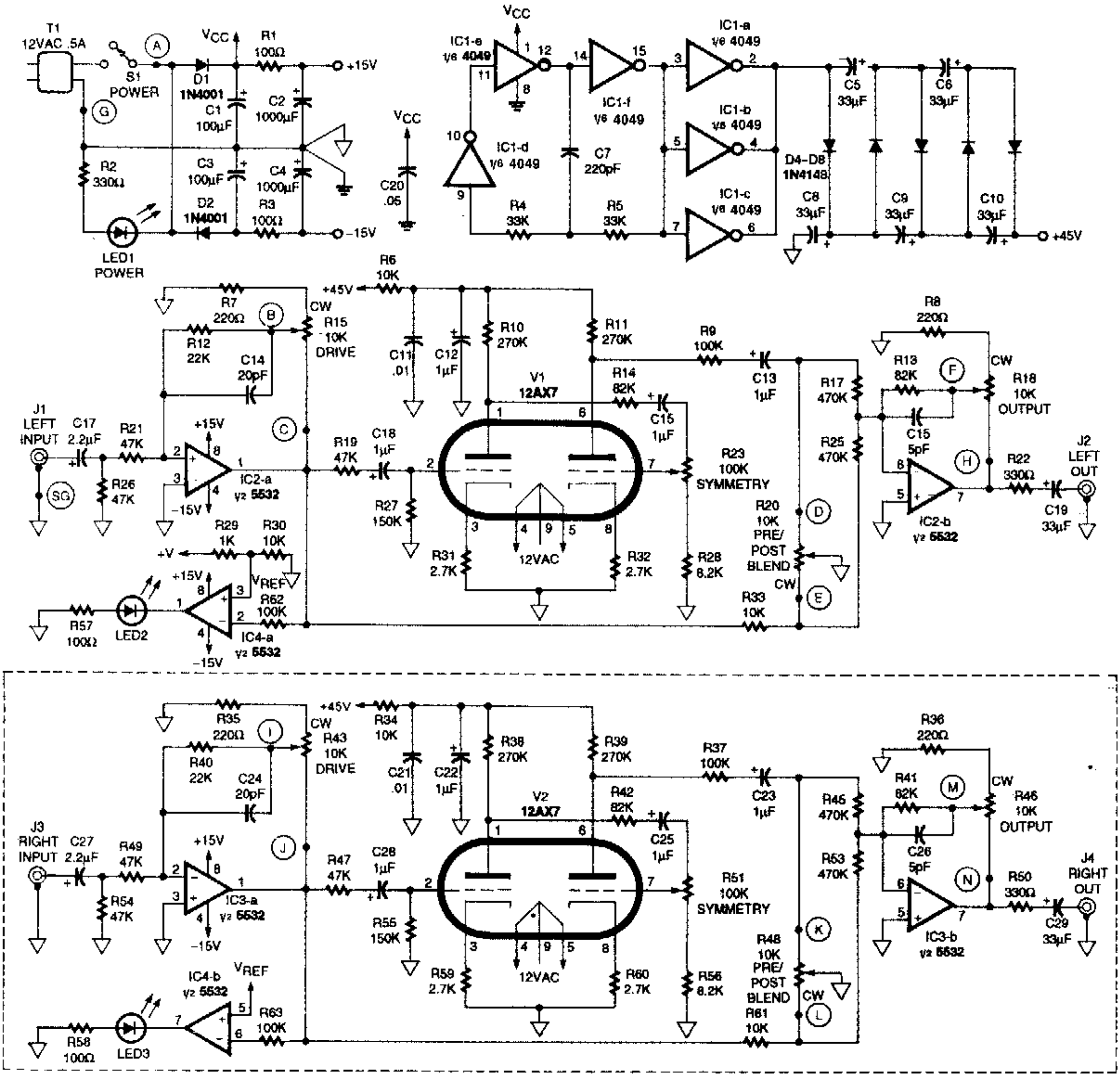
The front-end of this circuit is a bank of selectable bandpass filters. The output of the filter banks are shunted to ground via capacitor (C1) and PIN diode (D1). The PIN diode acts like an electronically variable resistor. The resistance across the diode's terminals is a function of the applied bias voltage. This voltage, hence the degree of attenuation of the RF signal, is proportional to the setting of potentiometer R1. The series resistor (R2) is used to limit the current when the diode is forward biased. This step is necessary because the diode has a very low resistance when a certain rather low potential is exceeded.

10

Audio Signal Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

VACUUM TUBE AUDIO AMPLIFIER



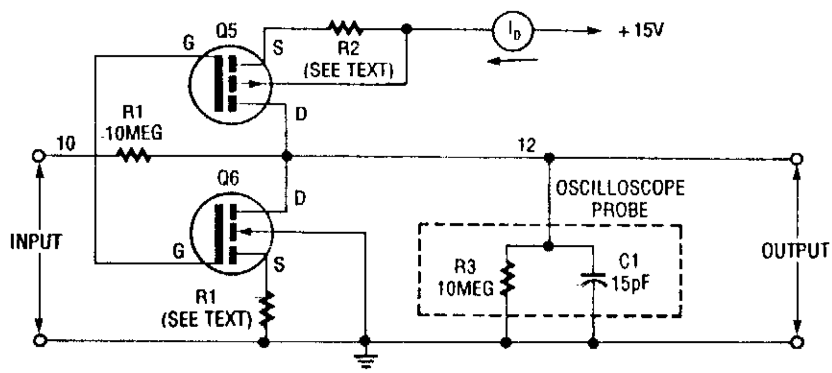
ELECTRONICS NOW *Fig. 10-1*

VACUUM TUBE AUDIO AMPLIFIER (*Cont.*)

This schematic is for a tubehead amplifier. The output from transformer T1 is positive half-wave ac rectified by D1 and filtered by C1, C2, and R1 for a +15-V supply. A −15-V supply is available from D2, C3, C4, and R3. The plate supply for the 12AX7 tubes is produced by a voltage multiplier.

Some listeners prefer the sound of a vacuum tube audio system. Although this is rather subjective and a personal preference, this circuit can be used to simulate the "tube sound" preferred by these listeners.

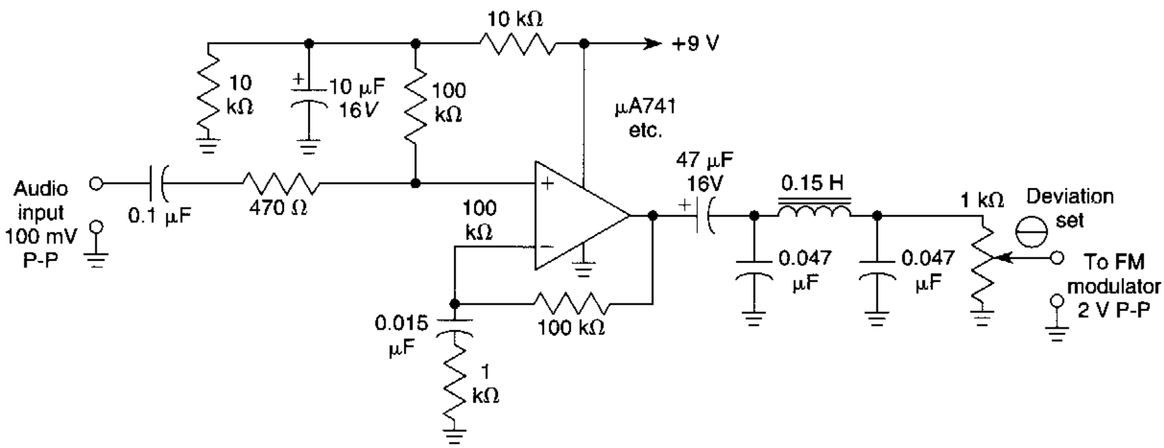
MICROPOWER LINEAR AMPLIFIER



ELECTRONICS NOW *Fig. 10-2*

This circuit, based on the inverter in the CD4007UB CMOS linear amplifier, shows a method for reducing drain current.

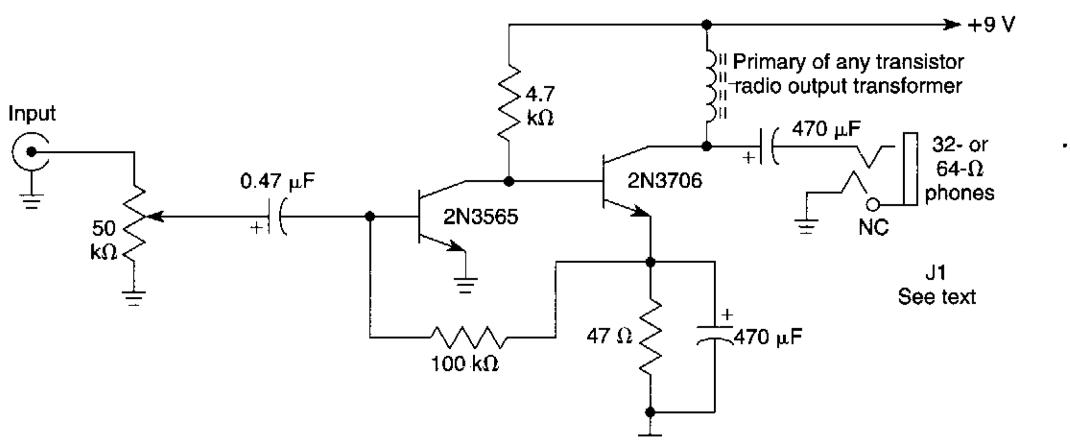
NB FM AUDIO AMPLIFIER



WILLIAM SHEETS *Fig. 10-3*

This audio system amplifies, limits, and filters an audio voice signal for use with an FM modulator or VCO. It has pre-emphasis of 6-dB/octave 300–3000 Hz. Almost any suitable op amp can be used.

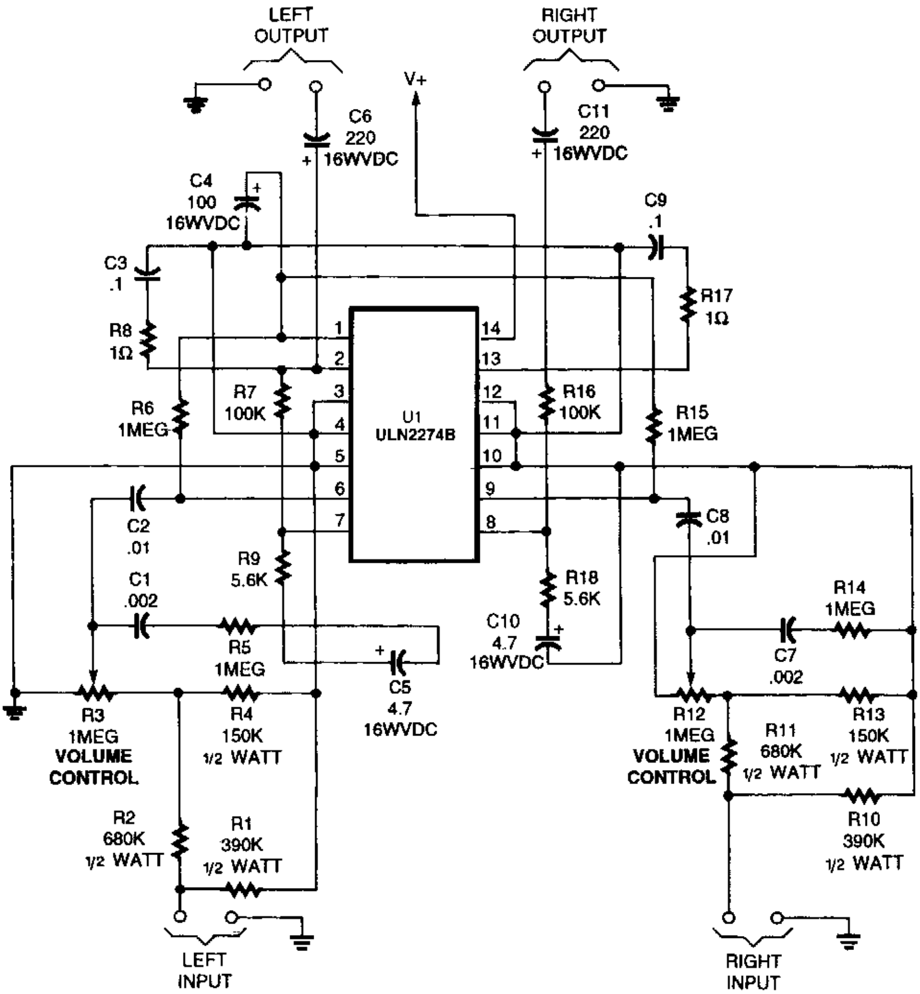
TWO-TRANSISTOR AUDIO AMPLIFIER



WILLIAM SHEETS *Fig. 10-4*

This is a general-purpose audio amplifier for driving a pair of stereo earphones in monaural mode. Two can be used for stereo. In this case, ground the center top of the earphone (sleeve of J1).

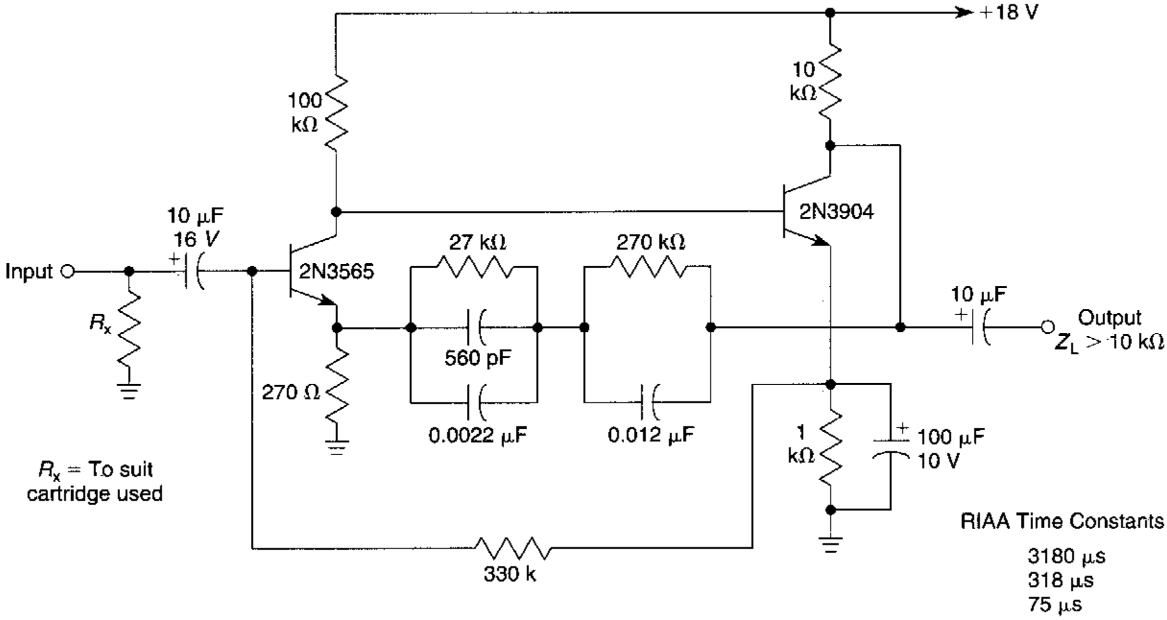
PERSONAL STEREO AUDIO AMP



POPULAR ELECTRONICS ***Fig. 10-5***

You can make your personal stereo do double duty as a small room stereo by adding this 2-watt amplifier.

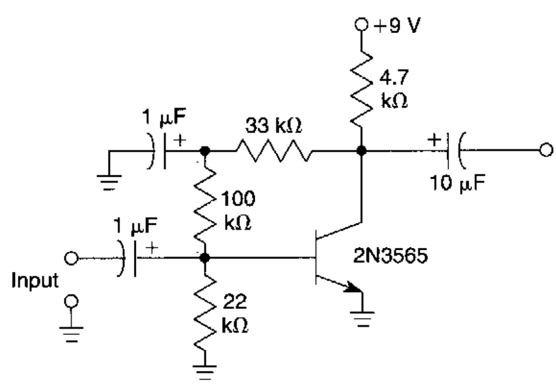
TRANSISTOR RIAA PREAMP FOR MAGNETIC PHONE CARTRIDGES



WILLIAM SHEETS *Fig. 10-6*

This two-transistor circuit has around 40 dB (midband) gain at 1 kHz. A magnetic cartridge is used as a source.

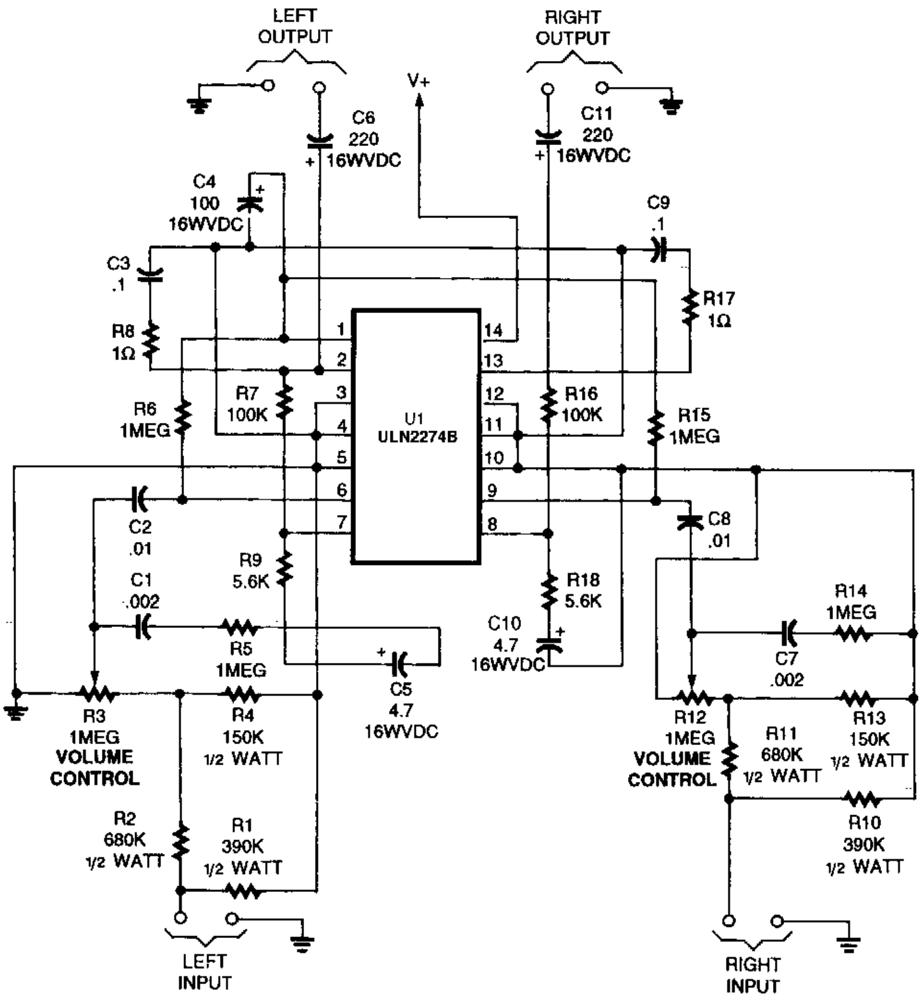
DYNAMIC MICROPHONE PREAMP



WILLIAM SHEETS *Fig. 10-7*

This preamplifier provides 40- to 43-dB gain when used with a low-impedance (<1 kΩ) dynamic microphone.

PERSONAL STEREO AUDIO AMP

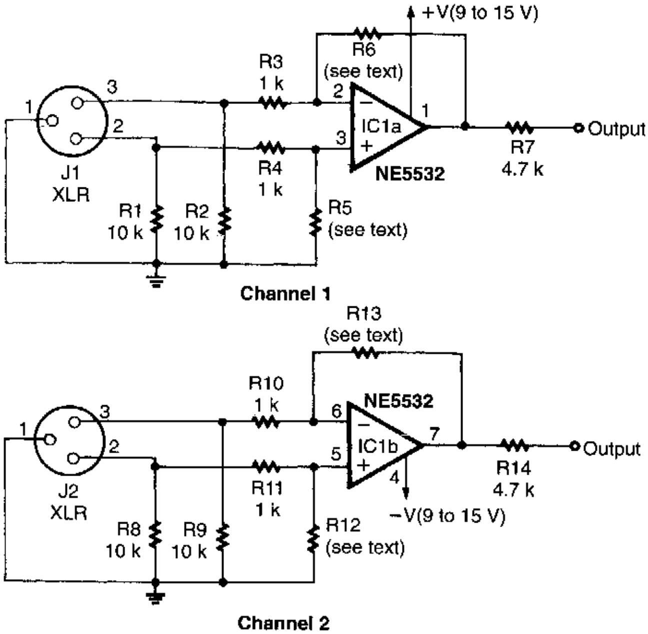


POPULAR ELECTRONICS

Fig. 10-5

You can make your personal stereo do double duty as a small room stereo by adding this 2-watt amplifier.

BALANCED MICROPHONE PREAMPLIFIER

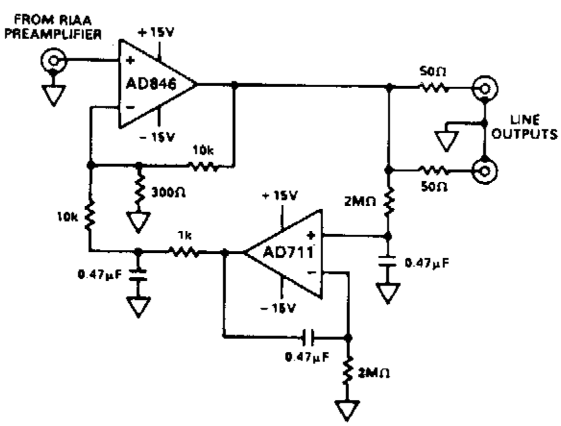


ELECTRONICS NOW

Fig. 10-8

A balanced input for microphones can solve hum and noise pickup problems. R6 and R13 should equal R5 and R12, respectively. Typical values would be 10 kΩ to 22 kΩ.

RIAA LINE AMPLIFIER/DRIVER

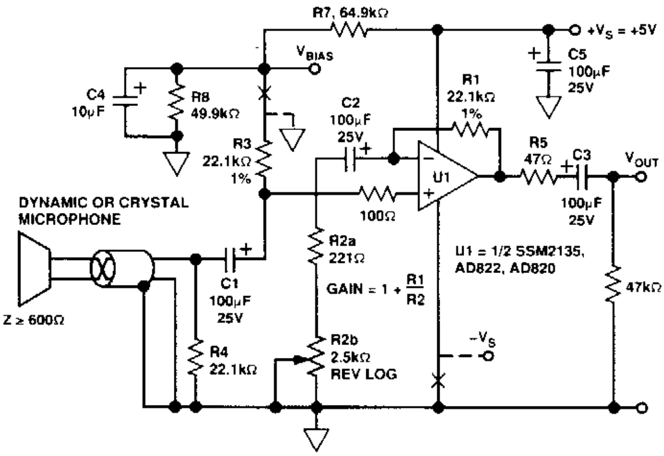


Two op amps by Analog Devices are used in this audio line amplifier, which is suitable for interfacing with an RIAA preamplifier.

ANALOG DEVICES

Fig. 10-9

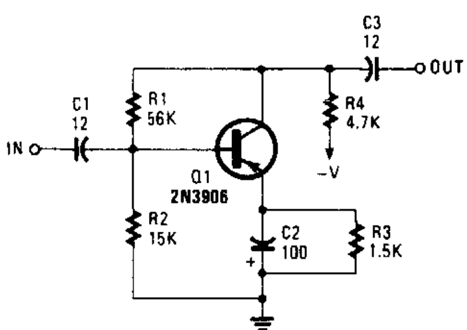
SINGLE-ENDED HI-*Z* MICROPHONE PREAMP



ANALOG DEVICES **Fig. 10-10**

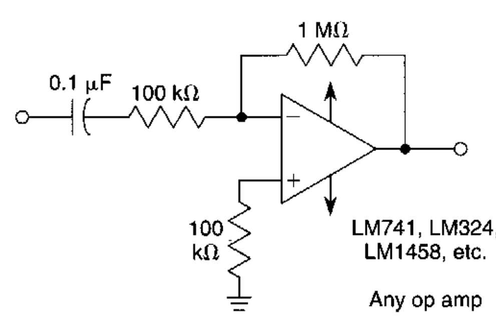
This low noise circuit works on a +5-V supply. Gain range is 20 to 40 dB and bandwidth is 20 kHz with the AD820. THD is 0.05% with 1 V RMS into a 2-kΩ load. Noise output with the input shorted is less than 200 μV.

LOW-LEVEL AUDIO AMPLIFIER



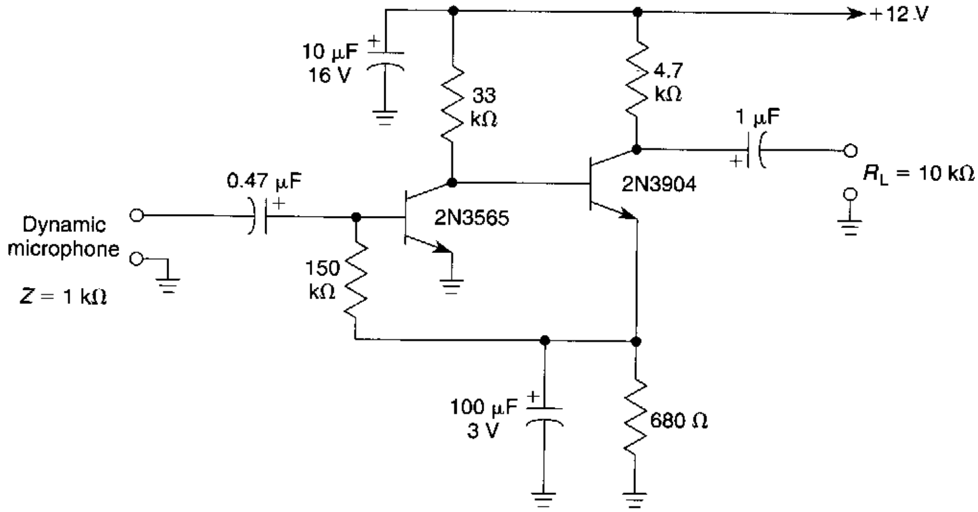
POPULAR ELECTRONICS **Fig. 10-11**

SIMPLE 20-dB GAIN AUDIO AMPLIFIER



WILLIAM SHEETS **Fig. 10-12**

HIGH-GAIN DYNAMIC MICROPHONE PREAMPLIFIER

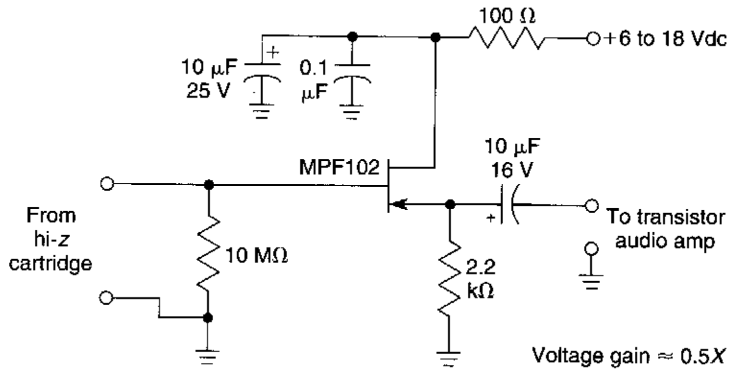


WILLIAM SHEETS

Fig. 10-13

This microphone preamplifier is capable of about 70 dB or more gain at audio frequencies. Its gain is approximately equal to the product of the h_{fe} of both transistors times the ratio of the load resistance to the input resistance of the preamp. As an approximation, these resistances are usually similar in value (≈2 to 5 kΩ) for most applications, so this ratio can be taken as unity.

FET PHONO CARTRIDGE PREAMP

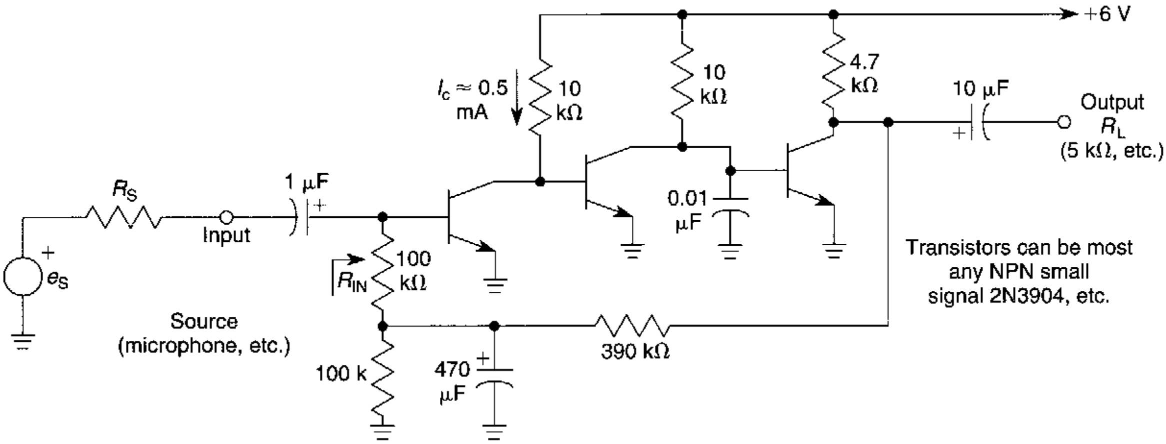


WILLIAM SHEETS

Fig. 10-14

A high-Z phono cartridge can be matched to a low-Z amplifier with this circuit. The FET provides a current gain of over 1000× and a voltage gain of about 0.5×.

SIMPLE HIGH-GAIN AUDIO AMPLIFIER

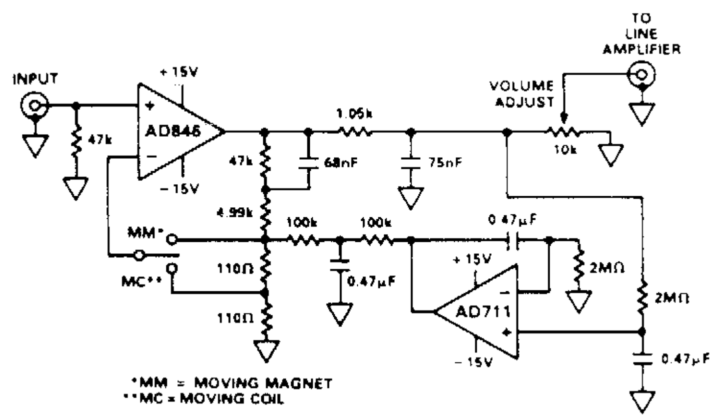


WILLIAM SHEETS **Fig. 10-15**

This amplifier has a very high gain in the audio range and is approximately the product of the current gains of the three transistors multiplied by the ratio of R_L to $(R_{IN} + R_S)$. R_{IN} is approximately to:

$$\frac{(\beta_{Q1} + 1)(26)}{I_{EQ1}}$$

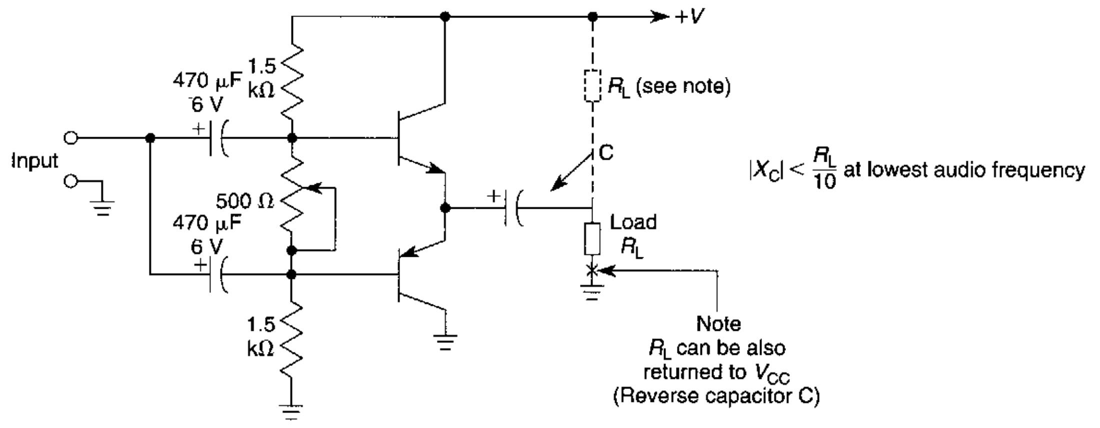
RIAA PREAMPLIFIER



ANALOG DEVICES **Fig. 10-16**

This preamp for RIAA phone use uses two op amps by Analog Devices. A switch selects compensation for moving magnet or moving coil pickups.

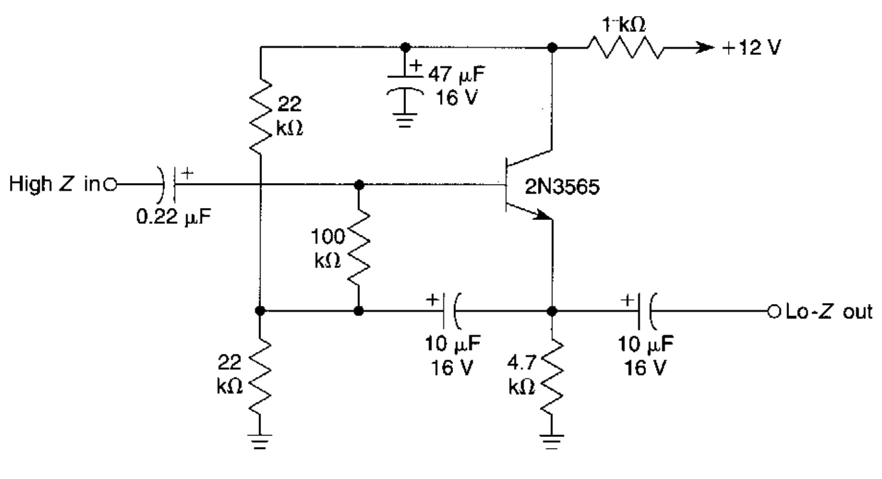
BASIC COMPLEMENTARY CLASS-AB SINGLE-SUPPLY AMPLIFIER



WILLIAM SHEETS

Fig. 10-17

HIGH-IMPEDANCE MICROPHONE INPUT CIRCUIT

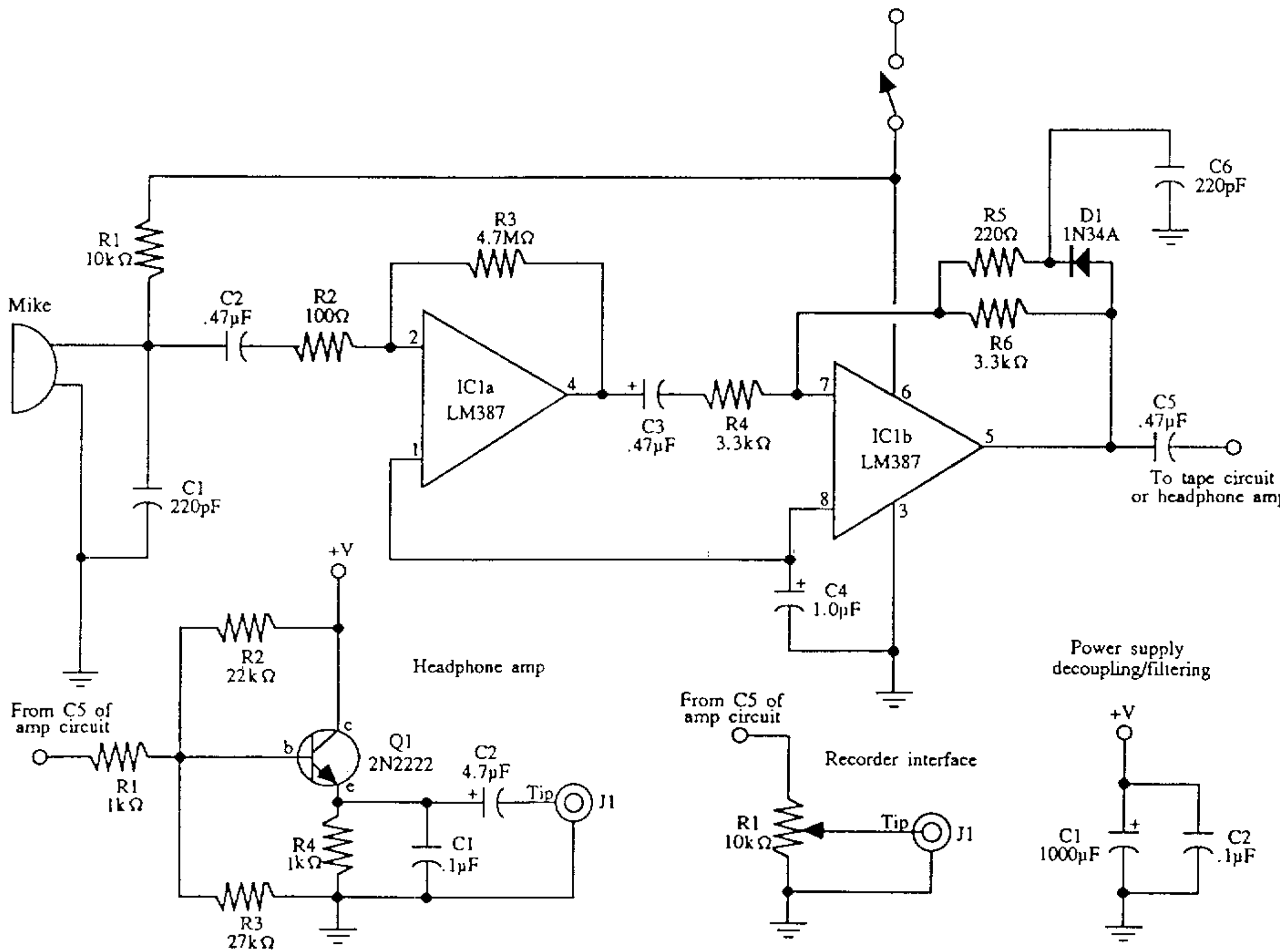


WILLIAM SHEETS

Fig. 10-18

This input circuit will enable use of a high-impedance microphone where a low-impedance microphone would be needed.

ELECTRONIC-EAR LOW-NOISE AUDIO AMPLIFIER (FOR PAROBOLIC DISH MIKES)



MCGRAW-HILL

Fig. 10-19

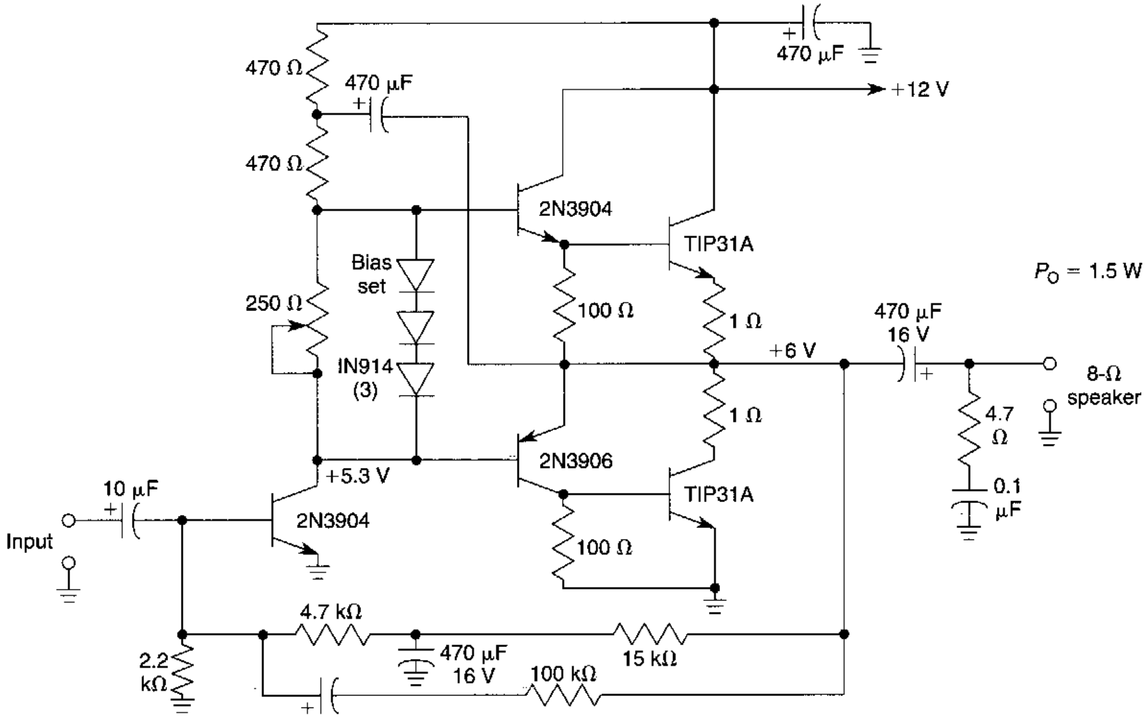
Use this circuit with a parabolic reflector microphone for eavesdropping on distant sounds.

11

Audio Power Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

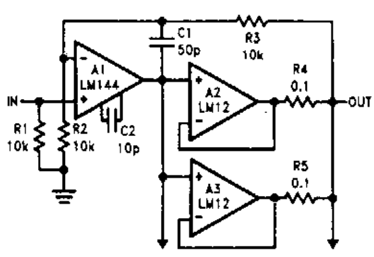
AUDIO POWER AMPLIFIER, 1.5 W, 12 V



WILLIAM SHEETS ***Fig. 11-1***

Although ICs have largely replaced circuits such as this, this circuit still finds use where the flexibility of a discrete device design is desirable. Parts are easy to obtain and the problem of IC obsolescence is eliminated. The TIP31A can be heatsinked to a small metal heatsink, if desired.

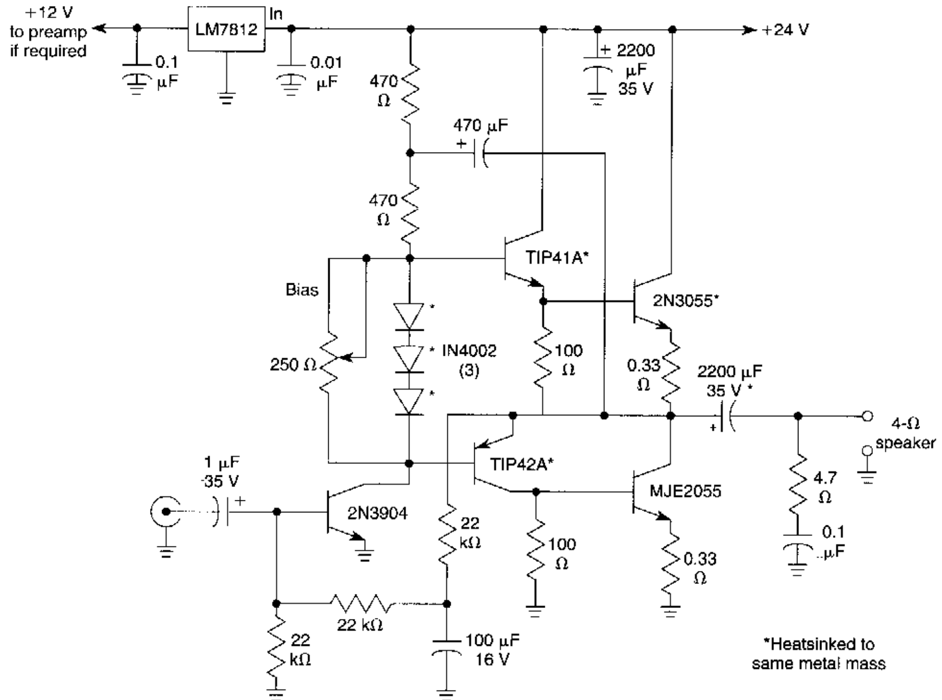
PARALLEL POWER OP AMPS



The power amplifiers, A2 and A3, are wired as followers and connected in parallel with the outputs coupled through equalization resistors.

NATIONAL SEMICONDUCTOR ***Fig. 11-2***

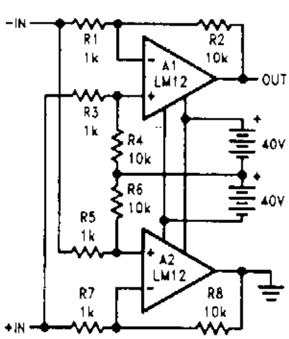
10-WATT AUDIO AMPLIFIER



WILLIAM SHEETS **Fig. 11-3**

This circuit is a general-purpose 10-W audio amplifier for moderate-power PA or modulator use in an AM transmitter. With higher voltages and a change in bias resistors, up to 30 W can be obtained.

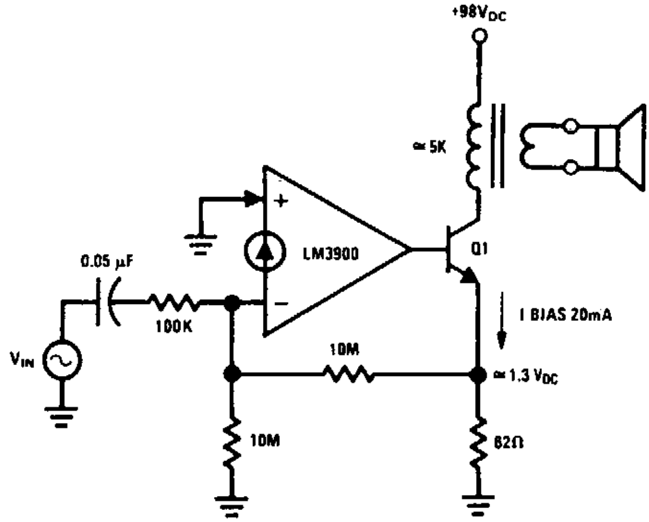
POWER BRIDGE AMPLIFIER WITH SINGLE-ENDED OUTPUT



Bridge amplifier with a single-ended output uses floating supply. Either input can be grounded.

NATIONAL SEMICONDUCTOR **Fig. 11-4**

LINE-OPERATED AUDIO AMPLIFIER

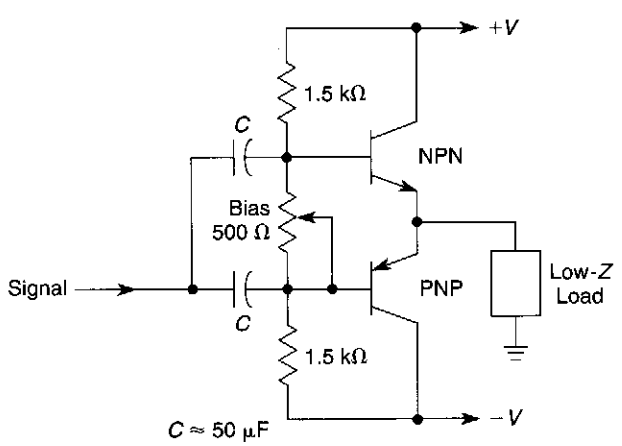


NATIONAL SEMICONDUCTOR

Fig. 11-5

An audio amplifier which operates off a +98-Vdc power supply (the rectified line voltage) is often used in consumer products. The external high-voltage transistor, Q1, is biased and controlled by the LM3900. The magnitude of the dc biasing voltage, which appears across the emitter resistor of Q1 is controlled by the resistor. The resistor is placed from the (–) input to ground.

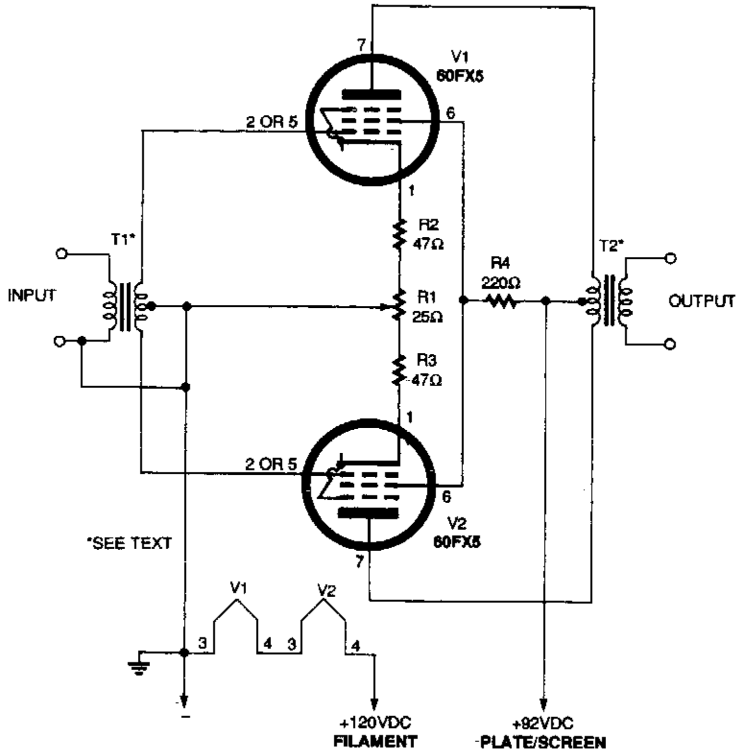
BASIC COMPLEMENTARY CLASS-AB POWER AMPLIFIER



WILLIAM SHEETS

Fig. 11-6

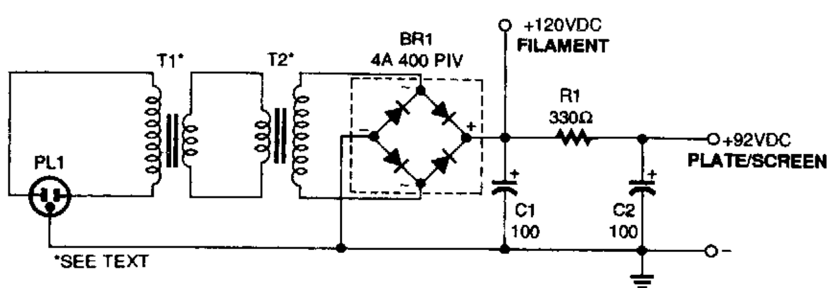
SIMPLE VACUUM TUBE AMPLIFIER



POPULAR ELECTRONICS *Fig. 11-7*

Using a pair of 60 FX5 tubes, direct operation from 120 Vac is possible. However, the use of a power supply with an isolation transformer is recommended. R1 is adjusted for equal voltages at pin 1 of V1 and V2. The power output is about 2 to 3 watts.

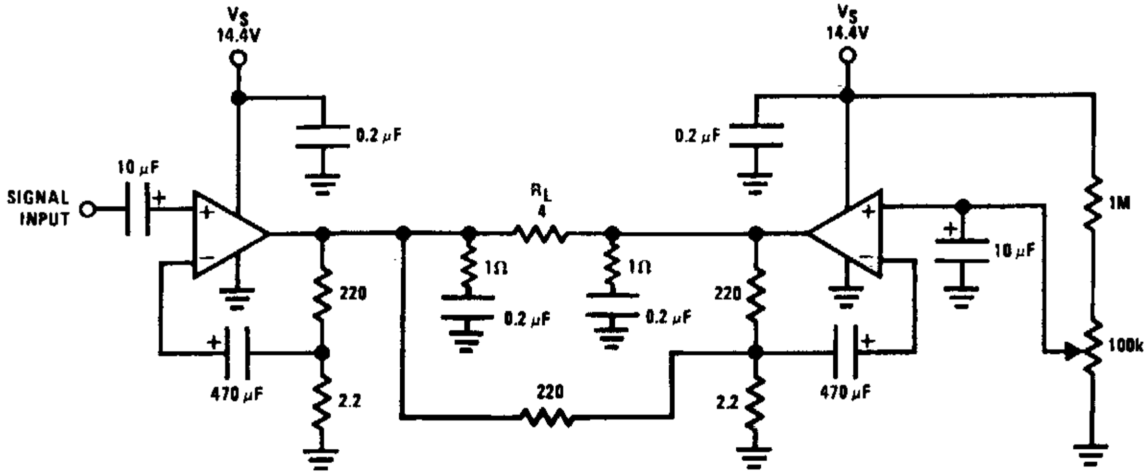
POWER SUPPLY FOR VACUUM TUBE AMPLIFIER



POPULAR ELECTRONICS *Fig. 11-8*

The power supply for the amplifier uses two low-voltage transformers connected back-to-back. The full-wave bridge rectifier, BR1 provides dc for the filaments, plates, and screens.

16-W BRIDGE AMPLIFIER

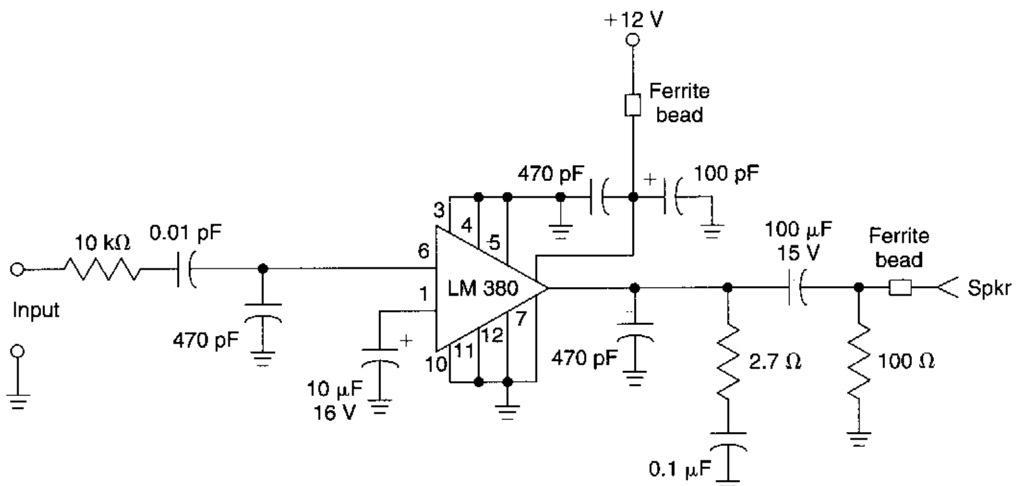


NATIONAL SEMICONDUCTOR

Fig. 11-9

This circuit delivers 16 W RMS audio into a 4-Ω load (R_L). The ICs are LM383s.

RFI-PROOF AUDIO POWER AMPLIFIER

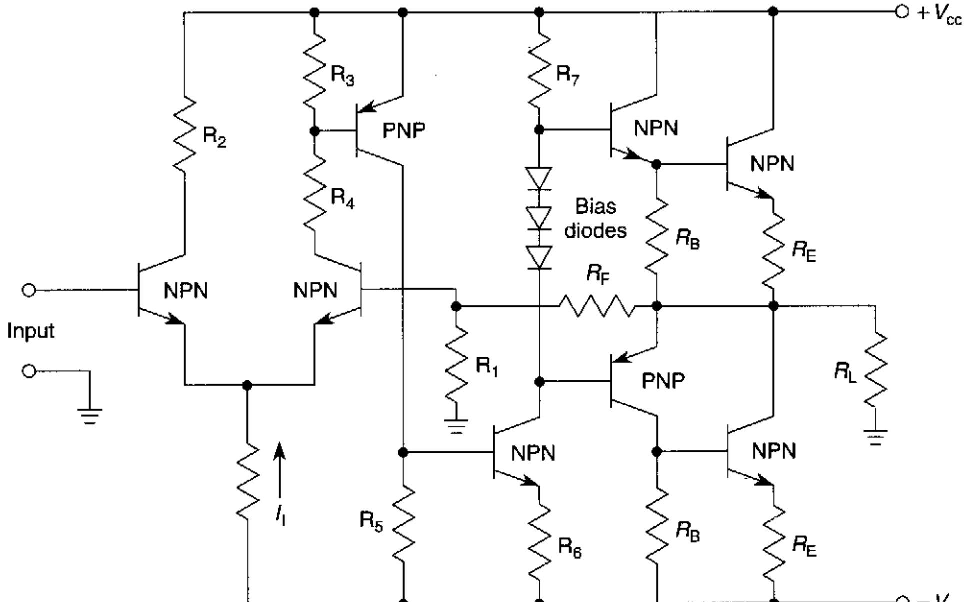


WILLIAM SHEETS

Fig. 11-10

This 1-watt audio amplifier was used in an FM repeater and proved to be immune to strong RF signal pickup. It functioned well in very strong RF fields.

BASIC QUASI-COMPLEMENTARY POWER AMPLIFIER WITH SPLIT POWER SUPPLIES

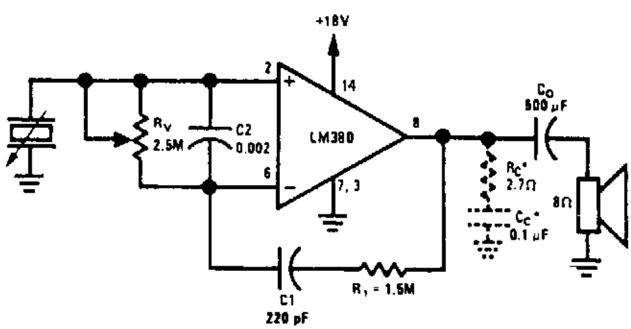


$$|GAIN| = \frac{R_F}{R_1}$$

WILLIAM SHEETS ***Fig. 11-11***

This is the basic circuit used in many audio power output stages where split supplies are used. This amplifier is inherently dc coupled and has high open loop gain and good dc stability if the feedback network is properly designed.

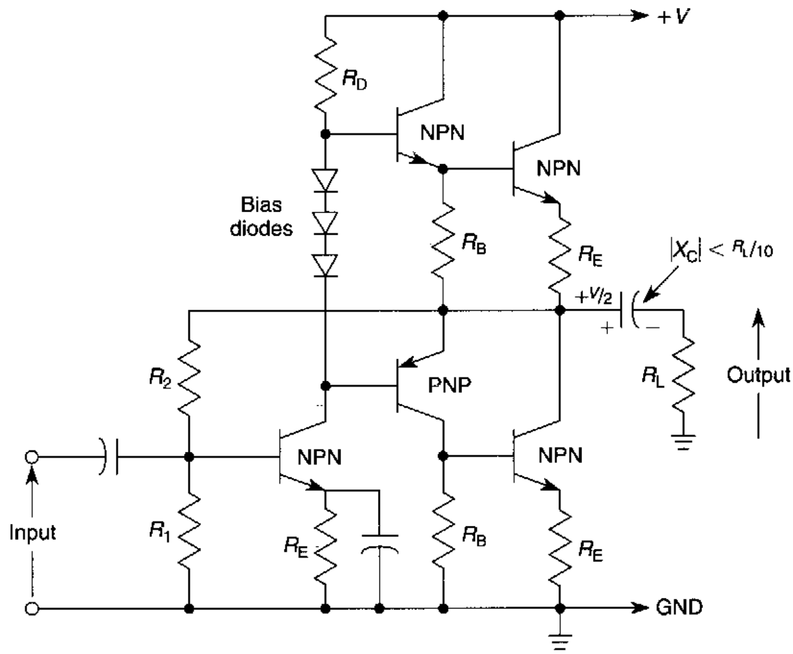
RIAA PHONO AMPLIFIER



NATIONAL SEMICONDUCTOR ***Fig. 11-12***

$$Mid\text{-}band\ gain = \frac{R_1 + 150\ \text{k}\Omega}{150\ \text{k}\Omega}$$

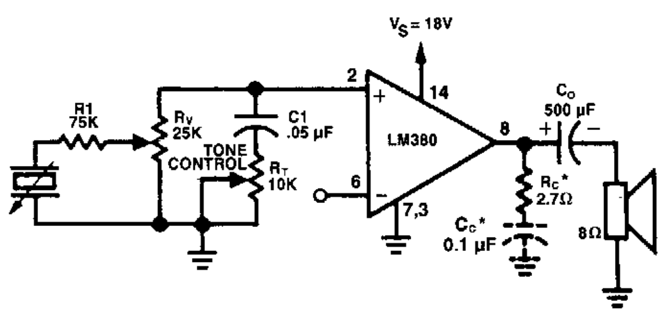
BASIC QUASI-COMPLEMENTARY POWER AMPLIFIER CIRCUIT



WILLIAM SHEETS

Fig. 11-13

PHONO AMP

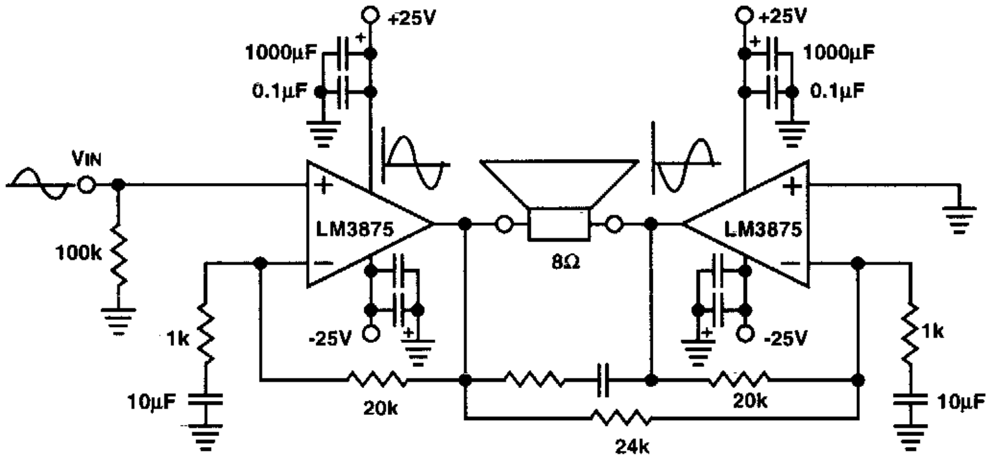


NATIONAL SEMICONDUCTOR

Fig. 11-14

The figure shows the LM380 with a voltage-divider volume control and high-frequency roll-off tone control.

80-WATT IC AUDIO AMPLIFIER

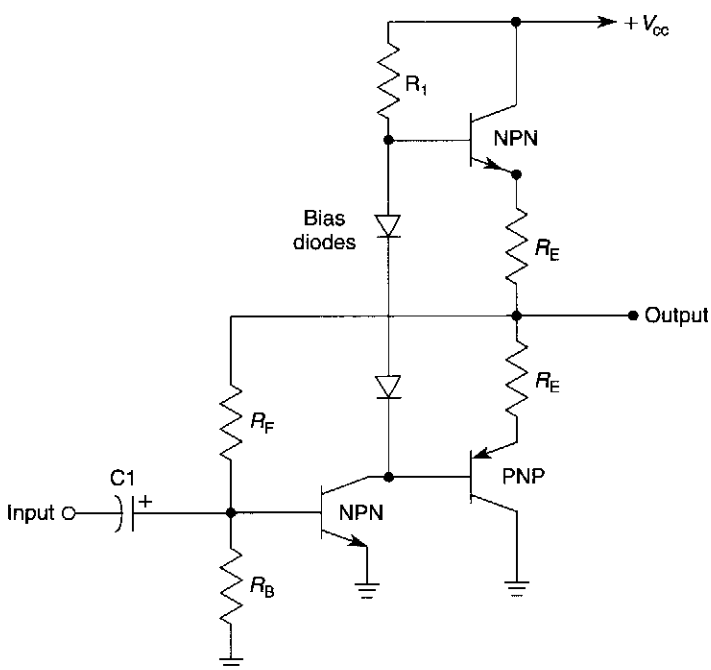


NATIONAL SEMICONDUCTOR

Fig. 11-15

This audio power amp will deliver 80 W of audio into an 8-Ω load. The LM3875 IC devices should be suitably heatsinked. Note that the amplifier is a bridged circuit, with both speaker leads "hot."

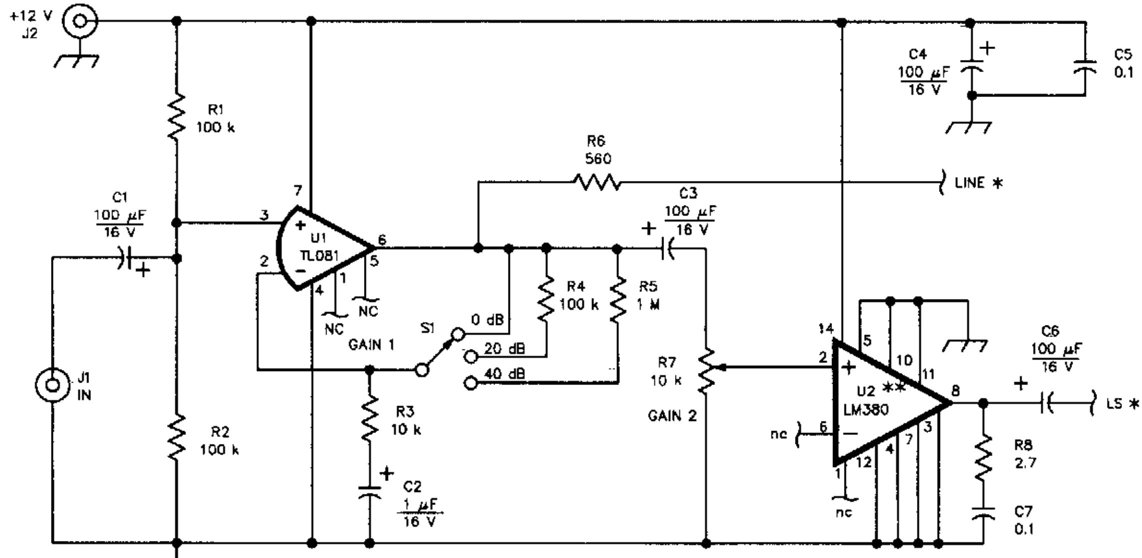
BASIC COMPLEMENTARY POWER AMPLIFIER CIRCUIT



WILLIAM SHEETS

Fig. 11-16

GENERAL-PURPOSE AF AMPLIFIER



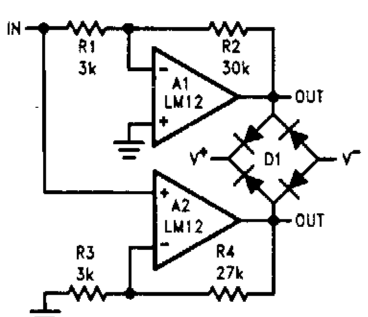
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.
IC Pins not shown are unused.
nc=Not connected.
*=See text.
**=Heat sink (see text).

Supply Voltage: 12 to 14 V
Input impedance: 50 kΩ (nominal)
Output impedance (LS[speaker] terminal): 8 Ω
Output impedance (LINE terminal): 560 Ω
Gain: Up to 70 dB, continuously variable
Bandwidth: 16 Hz to 30 kHz, minimum

QST ***Fig. 11-17***

Schematic of the general-purpose AF amplifier. All resistors are ¼-W, 5%-tolerance carbon-composition or metal-film units. Equivalent parts can be substituted. General-purpose IC replacements are shown in parentheses.

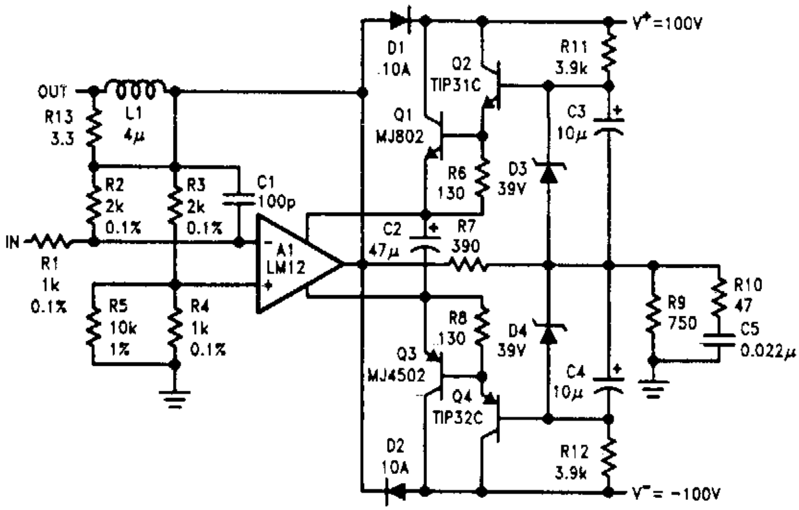
BRIDGE CONNECTION OF TWO POWER OP AMPS



These bridge connections provide differential outputs that approach twice the total supply voltage. Diode bridge clamps output to the supplies.

NATIONAL SEMICONDUCTOR ***Fig. 11-18***

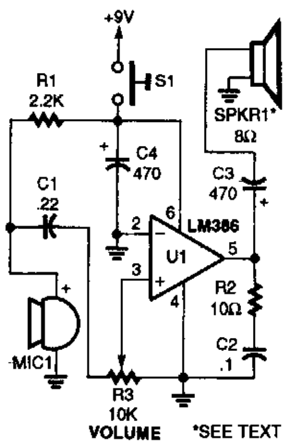
90-V 10-A HIGH-POWER AMPLIFIER



NATIONAL SEMICONDUCTOR ***Fig. 11-19***

This amplifier can drive ±90 V at 10 A, more than twice the output swing of the LM12. The IC provides current and power limiting for the discrete transistors.

MINI-MEGAPHONE



POPULAR ELECTRONICS ***Fig. 11-20***

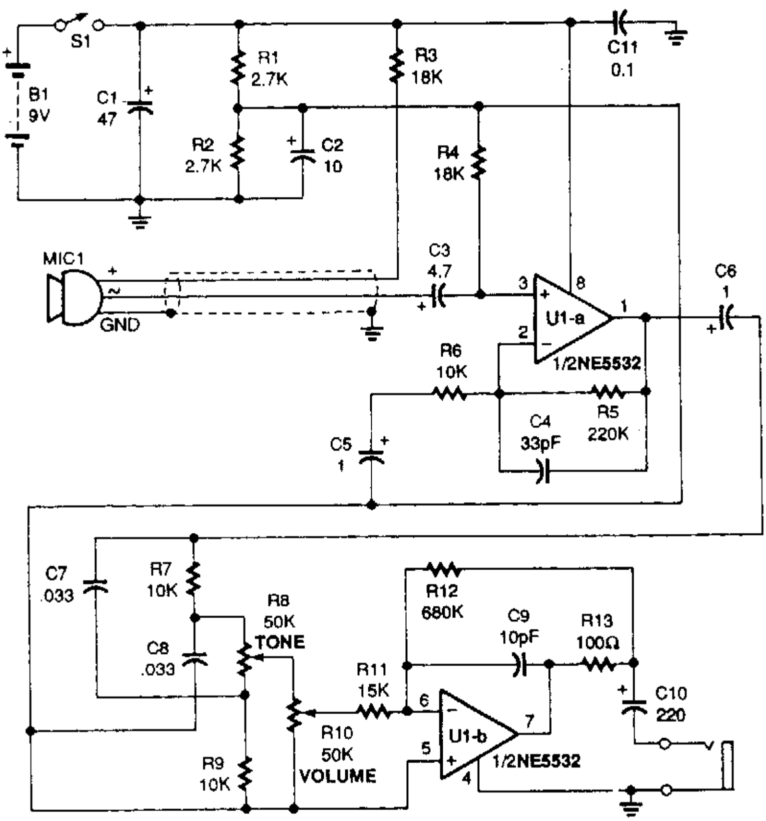
The Mini-Megaphone is comprised of an electret microphone (MIC1), and LM386 low-voltage audio-power amplifier (U1), a horn speaker (SPKR1), and a few other components.

12

Automotive Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC AUTO STETHOSCOPE



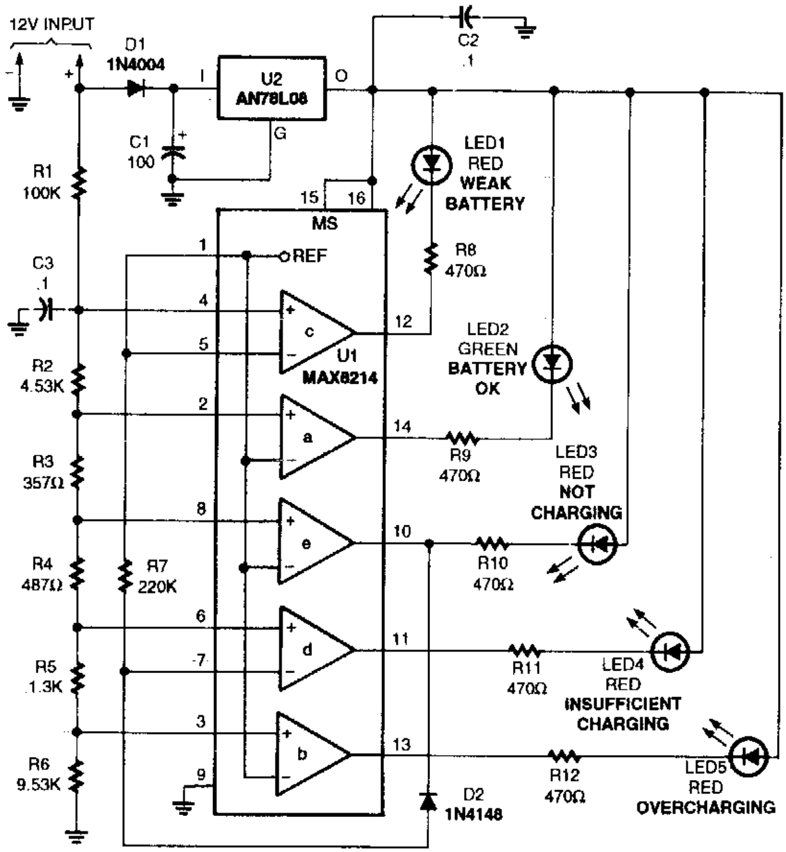
POPULAR ELECTRONICS *Fig. 12-1*

The heart of the Stethoscope is the NE5532 audio op amp, U1. That component directly drives low impedances and allows the use of headphones without adding another amplifier.

AUTOMOTIVE ELECTRICAL MONITOR

TABLE 1—AUTOMOTIVE ELECTRICAL FAULTS

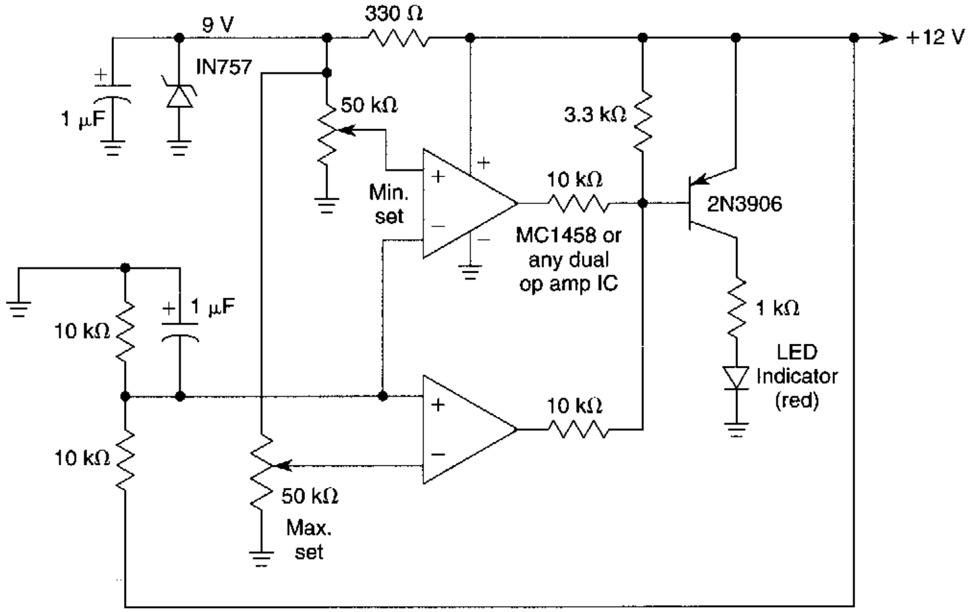
Condition	Normal Voltage	Possible Fault
Vehicle at rest	12.6 volts	<12.4 volts: bad cell or severely undercharged battery
Cranking	>9 volts	<9 volts: Weak battery
Idling	>12.8 volts	<12.8 volts: Not charging; bad alternator or wiring
Running minimum load	>13.4 volts	<13.4 volts: defective alternator or voltage regulator
Running minimum load	<15.2 volts	>15.2 volts: Overcharging; defective regulator
Running maximum load	>13.4 volts	<13.4 volts: alternator defective or belt slipping



POPULAR ELECTRONICS *Fig. 12-2*

The automotive electrical diagnostic system is built around a Maxim MAX8214ACPE five-stage voltage comparator, which contains a built-in 1.25-volt precision reference, and on-board logic that allows the outputs of two of the comparators to be inverted.

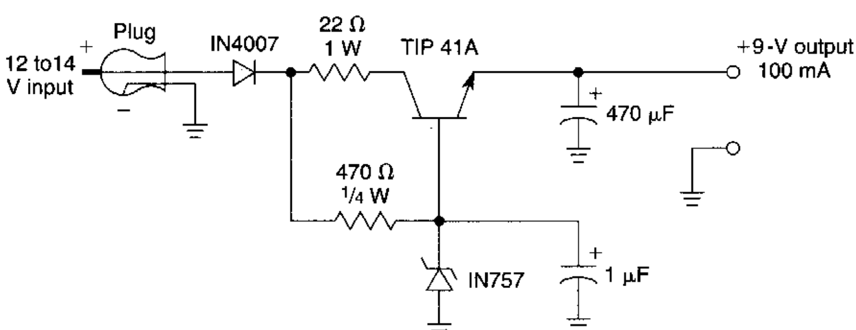
CAR ALTERNATOR MONITOR (IDIOT LIGHT)



WILLIAM SHEETS ***Fig. 12-3***

A window comparator is used to detect a too-low or a too-high system voltage. The minimum and maximum settings are set with two 50-kΩ pots, as desired.

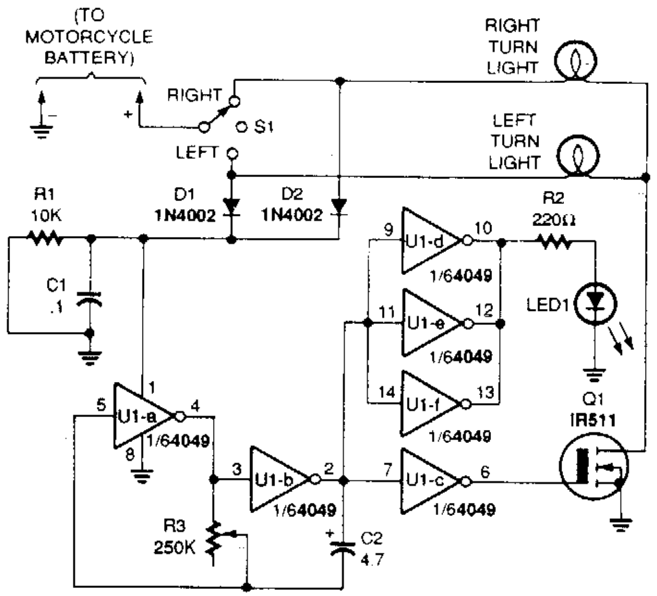
CIGARETTE LIGHTER 9-V ADAPTER



WILLIAM SHEETS ***Fig. 12-4***

A simple way to provide +9 V at 100 mA from a 12-V auto source. Applications include small radios, cassettes, etc.

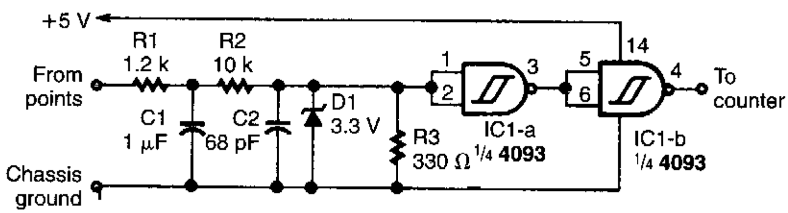
MOTORCYCLE TURN-SIGNAL SYSTEM



POPULAR ELECTRONICS *Fig. 12-5*

Tired of making hand signals? Build this simple turn-signal system and keep your hands on the handlebars.

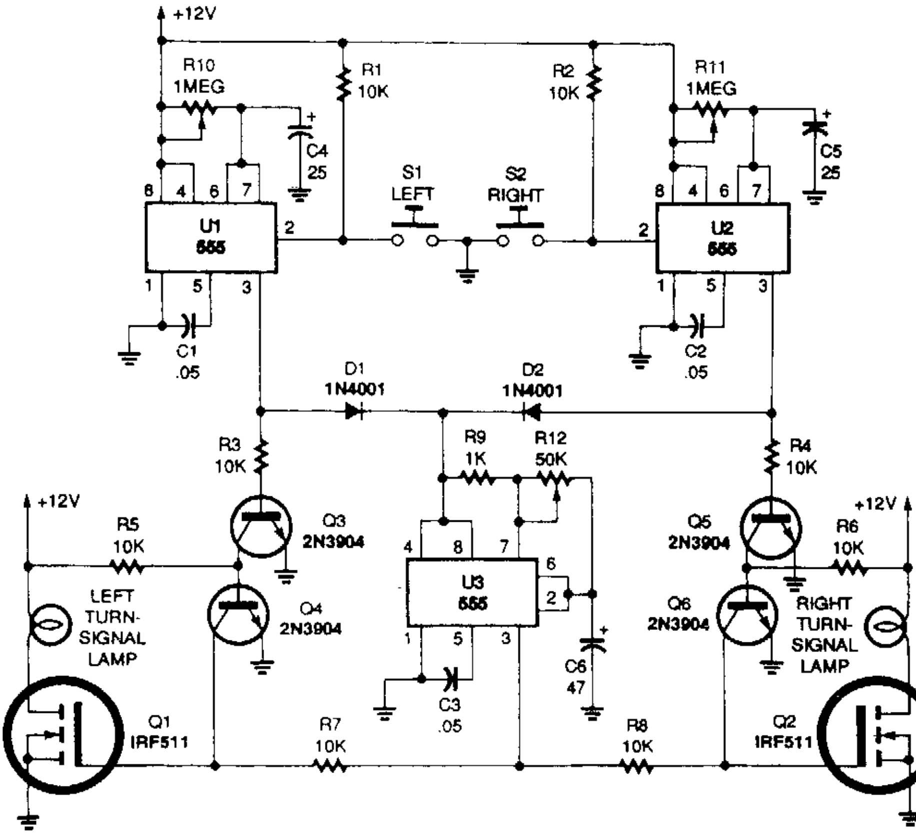
TACHOMETER SIGNAL-CONDITIONING CIRCUIT



ELECTRONICS NOW *Fig. 12-6*

This circuit, for use with auto tachometers, cleans up the ragged distribution waveform before it is sent to pulse counter circuits.

SMART TURN SIGNAL FOR AUTOS AND MOTORCYCLES



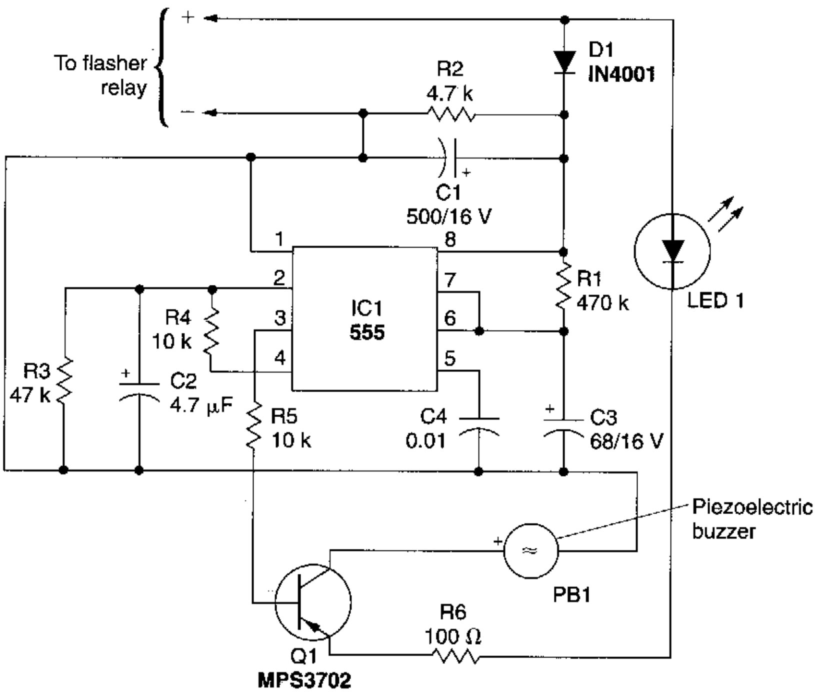
POPULAR ELECTRONICS *Fig. 12-7*

Momentarily pressing S1 starts the left on-time timer and produces a positive output at pin 3 of U1. Power for the on/off signal timer, U3, is supplied through D1.

Also, a positive bias is supplied from U1's output to the base of Q3, turning it on and turning Q4 off. Unclamped Q1 turns the left turn-signal lamp on and off at that same low-frequency rate. Because U2 is not activated, its output at pin 3 is low, keeping Q5 off. With Q5 turned off, Q6 is on, clamping the gate of Q2 to ground and keeping it from responding and supplying an output for the right turn-signal lamp. The left turn signal continues to operate until the U1 timer circuit times out; the right turn signal operates in a similar manner, with U2 setting its operating time.

Potentiometer R10 sets the running time for the left turn signal and R11 sets that for the right turn signal.

TURN SIGNAL ALARM



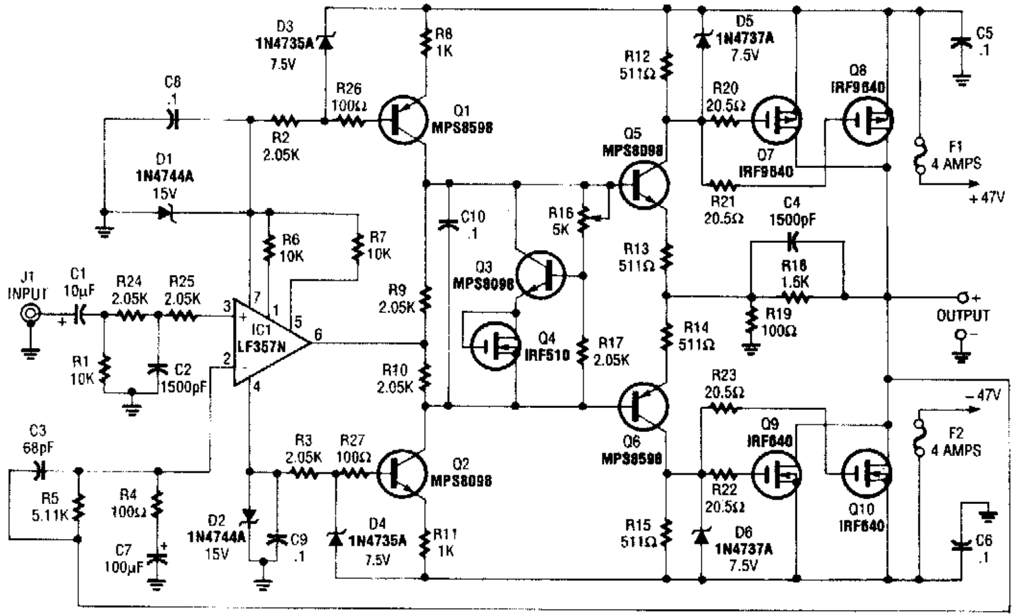
RADIO-ELECTRONICS *Fig. 12-8*

This circuit can be used to tell the driver of a vehicle when his or her turn signal has been left on for too long. The circuit consists of IC1, a 555 timer; transistor Q1, and MPS3702 PNP preamp/driver; PB1, a piezoelectric buzzer; along with an assortment of resistors, capacitors, and diodes. The 555 is connected in the monostable mode, requiring only a momentary negative pulse at pin 2 to trigger the timing cycle.

Power for the circuit is picked off the flasher relay and applied to IC1, pin 8, provided by an initially discharged capacitor, C2. After the initial triggering, the voltage across C2 rises as it becomes charged through R4, a 10-kΩ resistor. This prevents subsequent interference with the delay function caused by false triggering.

Capacitor C3 and resistor R1 determine the delay. With the component values shown, a delay of about one minute will be provided before the intermittent tweet sound generated by the circuit begins. If higher values are used for C2 and R1, a longer delay time will result. The light-emitting diode, LED1, provides a voltage drop to assure complete transistor blocking during the off periods of the flasher. Alternatively, two diodes in series can be used.

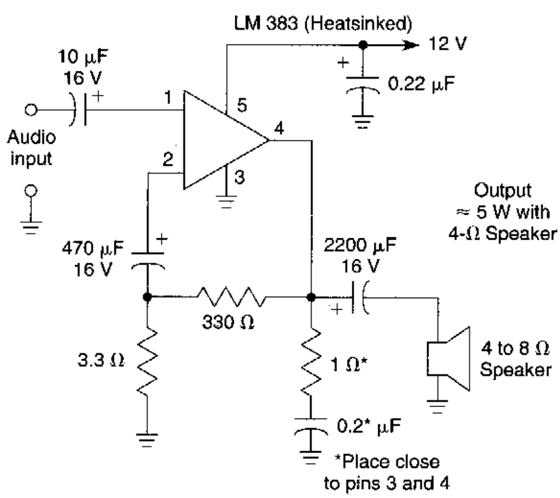
HIGH-POWER AUDIO AMP FOR AUTOMOTIVE INSTALLATION



RADIO-ELECTRONICS *Fig. 12-9*

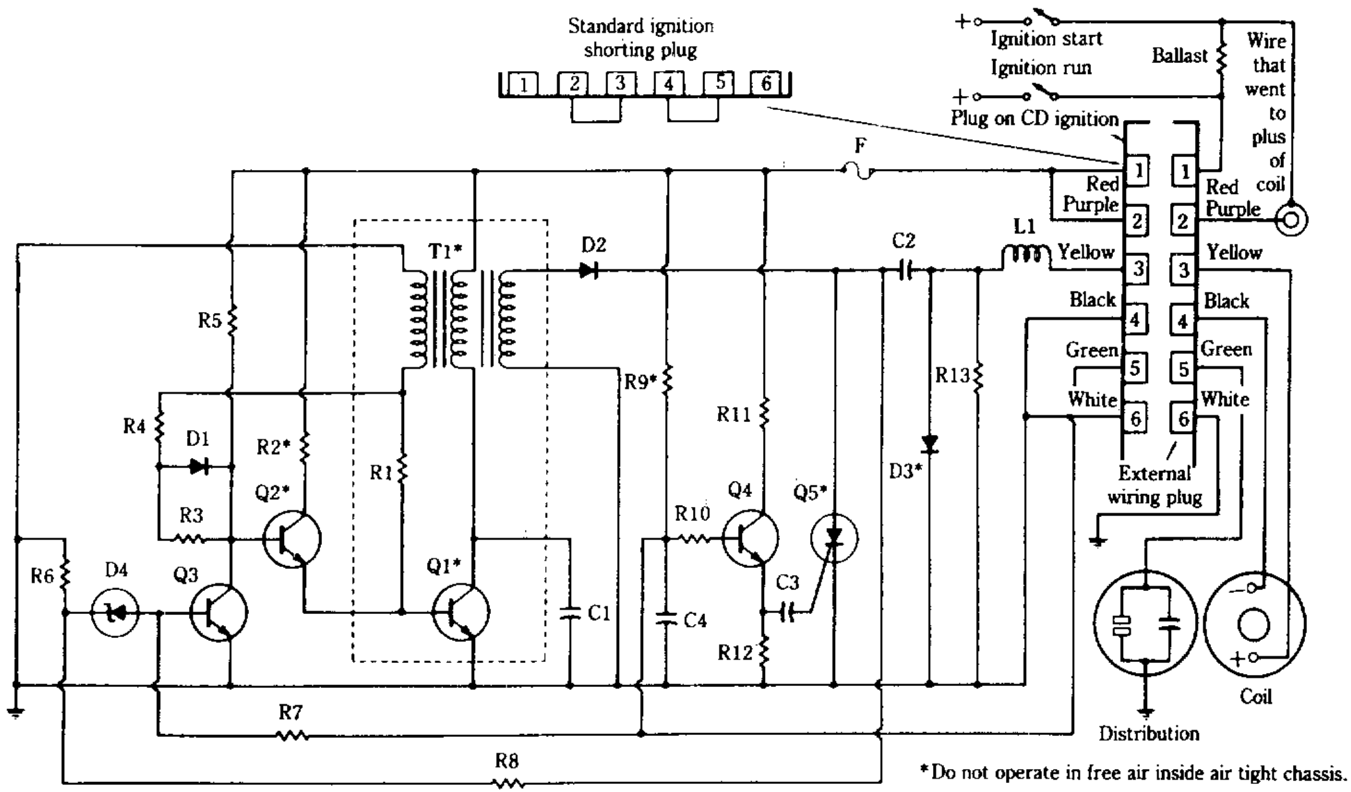
Two of these audio amplifiers can be used to make a stereo amplifier 200 W per channel. IRF640 and IRF9640 power MOSFETs are used to drive the output load, which might be 4 or 8 Ω. Response is 12 Hz to 45 kHz (–3 dB), THD <0.1%. Power is supplied by a switching-type power supply, which is external to the amplifier (±47 V). About 600 W total power (peak) is needed.

HIGH-POWER 12-V IC AUTO AMPLIFIER



WILLIAM SHEETS *Fig. 12-10*

CAPACITOR DISCHARGE IGNITION SYSTEM



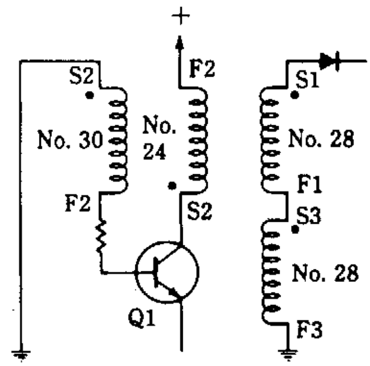
MCGRAW-HILL

Fig. 12-11

This ignition system charges a capacitor (C2) to 350 V and discharges it through the ignition coil.

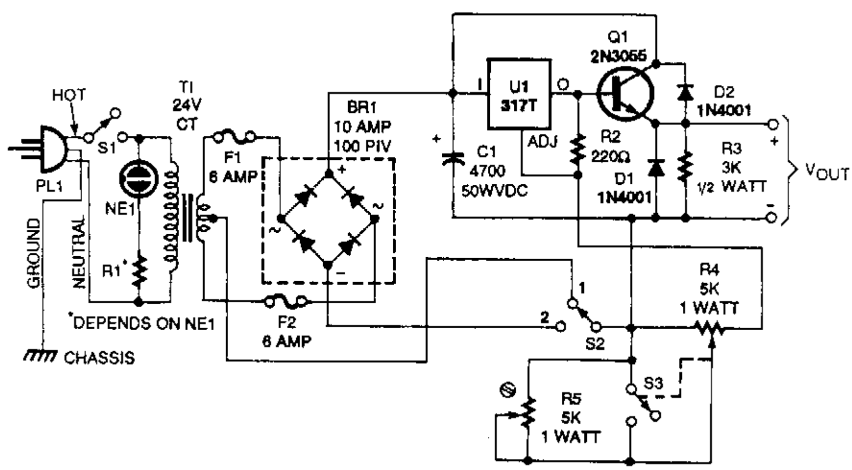
A Parts list

Q1—2N3055
Q2—2N3053
Q3—2N3241
Q4—2N3241
Q5—RCA 40657
D1—1N3193
D2—1N3195
D3—1N1763A
D4—12 V, ¼ W
C1—0.25 μF, 200 V
C2—1 μF, 400 V
C3—1 μF, 25 V
C4—0.25 μF, 25 V
F—5A
L1—10 μH, 100 Turns of No. 28 Wire Wound on a 2-W Resistor (100 Ohms or More)
R1—1000 ohms, ½ W
R2—35 ohms, 5 W
R3—22,000 ohms, ½ W
R4—1000 ohms, ½ W
R5—18,000 ohms, ½ W
R6—15,000 ohms, ½ W
R7—8200 ohms, ½ W
R8—0.39 megohm, ½ W
R9—220 ohms, 1 W
R10—1000 ohms, ½ W
R11—68 ohms, ½ W
R12—4700 ohms, ½ W
R13—27,000 ohms, ½ W



Details of inverter transformer

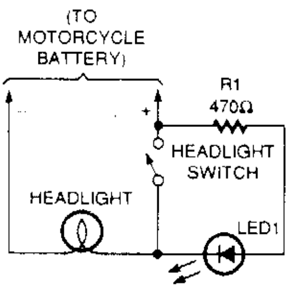
CAR AUDIO POWER SUPPLY



POPULAR ELECTRONICS *Fig. 12-12*

This supply has a variable output voltage feature and a dual voltage switch, S2. Q1 should be adequately heatsinked.

MOTORCYCLE HEADLIGHT MONITOR

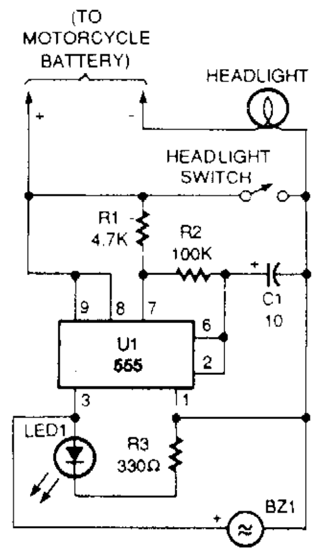


POPULAR ELECTRONICS ***Fig. 12-13***

The headlight on most newer bikes is keyed on with the ignition switch to guarantee that you are never underway without your headlight being on. However, many older bikes have a factory headlight switch, and a growing number of the newer bikes are owner-modified in the same way.

A simple headlight monitor circuit consists of just an LED and a current-limiting resistor wired across the headlight switch, as shown. When the ignition is on and the headlight switch is off, the LED will glow.

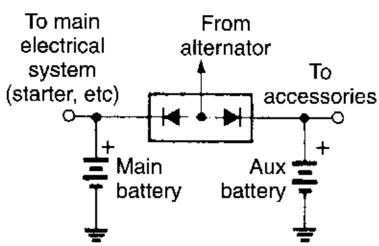
HEADLIGHT-OFF INDICATOR



POPULAR ELECTRONICS ***Fig. 12-14***

Increasing the value of R_2 or C_1 will lower the oscillator's frequency and decreasing one of those values will increase the frequency. The IC's output at pin 3 drives the LED through R3 and sends power to the piezo sounder. Use a bright LED so that you will be able to see it in the daytime.

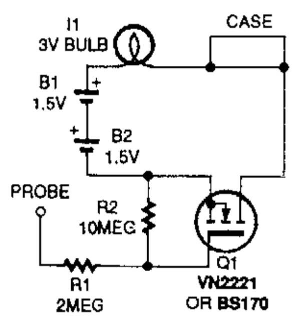
AUTO BATTERY ISOLATOR CIRCUIT



ELECTRONICS NOW ***Fig. 12-15***

The diodes ensure that current can flow in both batteries from the alternator, but the main battery can't feed the accessory system, nor vice versa.

AUTOMOTIVE HI-*Z* TEST LIGHT



POPULAR ELECTRONICS ***Fig. 12-16***

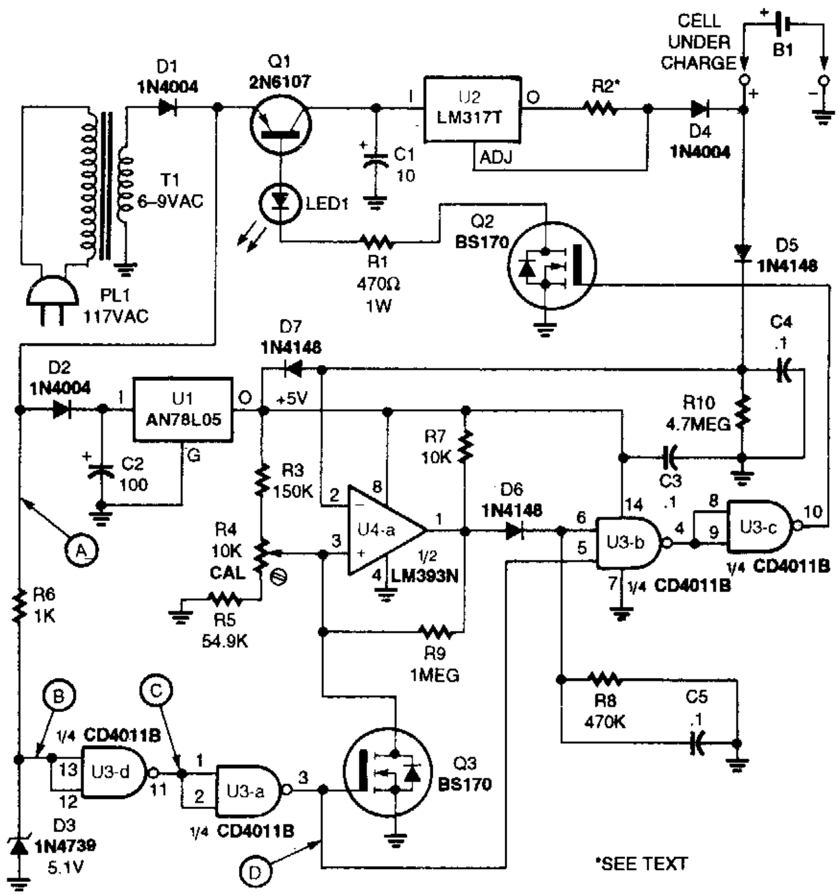
This test light has a high-input impedance and draws only 1 mA at 12 V. Q1 switches dc to a battery and lamp circuit.

13

Battery Charger Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

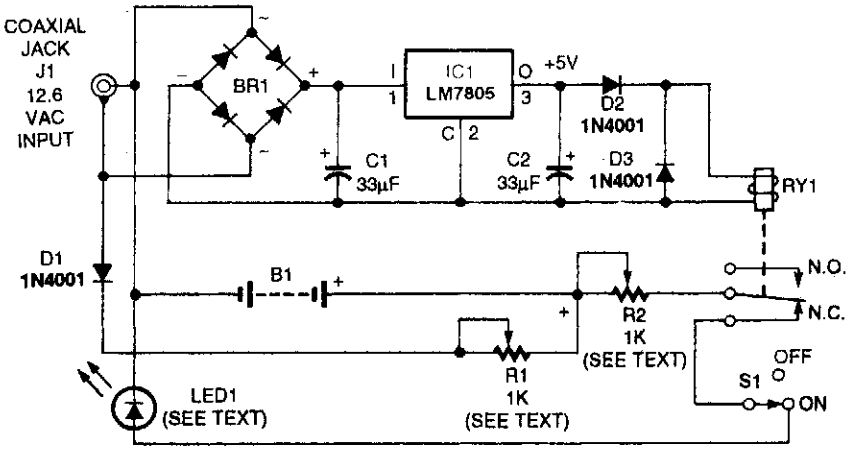
SMART BATTERY CHARGER



ELECTRONICS NOW *Fig. 13-1*

This charger will work with NiCad or the new rechargeable alkaline batteries. The Smartcharger is comprised mainly of four chips—an AN78L05 5-V, 100-mA regulator (U1), an LM317T 1-A adjustable-voltage regulator (U2), a CD4011BE quad 2-input NAND gate (U3), and an LN393N dual-voltage comparator (U4). The value and rating of R2 is selected as described in the text. R2 is selected for a 1.2-V drop across it at the charging current (3 Ω for 400 mA, 6 Ω for 200 mA).

RECHARGEABLE LED FLASHLIGHT

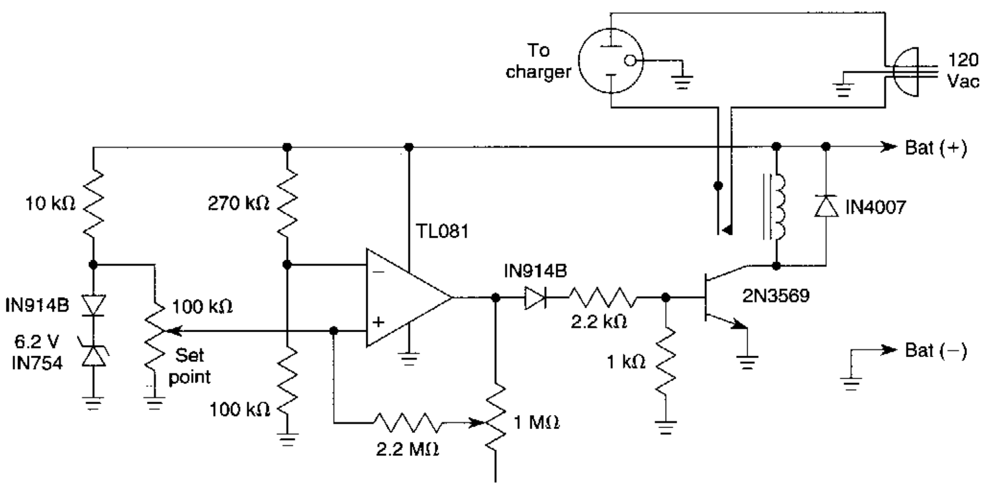


ELECTRONICS NOW

Fig. 13-2

This flashlight is useful for applications where night vision and/or darkness adaptation must be maintained. It uses an HLMP8150 T4 LED with a wavelength of 637 nm. This schematic is for the flashlight module. When the battery pack consisting of the four NiCad cells is fully charged (and there is no voltage at J1), 4.8 Vdc flows through trimmer potentiometer R2, the normally closed contact of relay RY1, and push-on/push-off power switch A1. Trimmer R2 limits the current flowing through LED1. Switch S1 can turn LED1 on and off when the battery is not being charged.

BATTERY CHARGER CONTROLLER

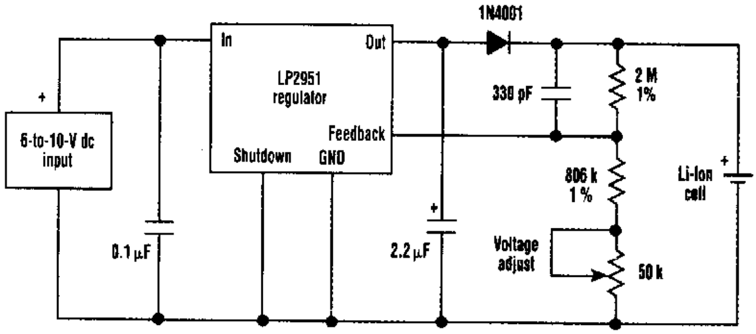


WILLIAM SHEETS

Fig. 13-3

When the battery voltage is low, the TL081 comparator produces a high output, turning on the 2N3569 relay driver. As the battery voltage approaches the set point, the relay driver is cut off, opening the 120-Vac supply.

SINGLE-CELL LITHIUM BATTERY CHARGER



ELECTRONIC DESIGN

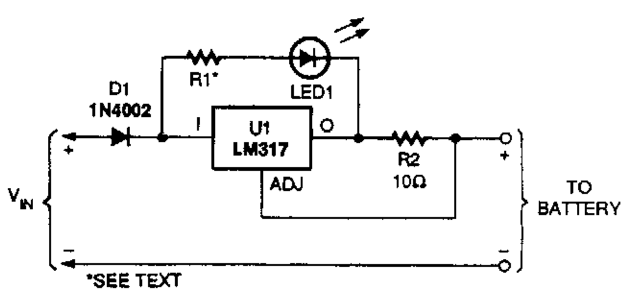
Fig. 13-4

An LP2951 regulator was chosen for this single lithium cell-charging circuit for its built-in current-limiting capability. In addition, the regulator's output voltage is extremely stable, which is a prerequisite for lithium battery charging. This figure details an example circuit designed to recharge a single cell. The required output set voltage was specified as 4.200 V (±0.025 V) with a maximum charging current of about 150 mA.

An LP2951 regulator was selected for two reasons. One is that its built-in current limiter holds the maximum current to 160 mA (typical). The other is because the output voltage can be very accurately set to 4.200 V, thanks to the regulator's stable internal bandgap reference.

The 1.23-V reference appears between the feedback pin and ground, which causes a precise current to flow in the output resistive-divider string. The amount of current flowing in these resistors determines (sets) the charger output voltage that appears across the battery terminals. Large-value resistors keep the battery drain below 2 µA when the dc input is removed (a customer requirement). A trimming potentiometer sets the output to 4.200 V. It must be adjusted when the battery isn't connected to the charger output. A blocking diode is required at the LP2951's output to prevent current from flowing out of the battery and back into the output when the dc-input source is removed. Because the diode is in series with the output, the minimum input-output voltage differential required for this circuit to operate is about 1.5 V.

BATTERY-CHARGING CURRENT LIMITER

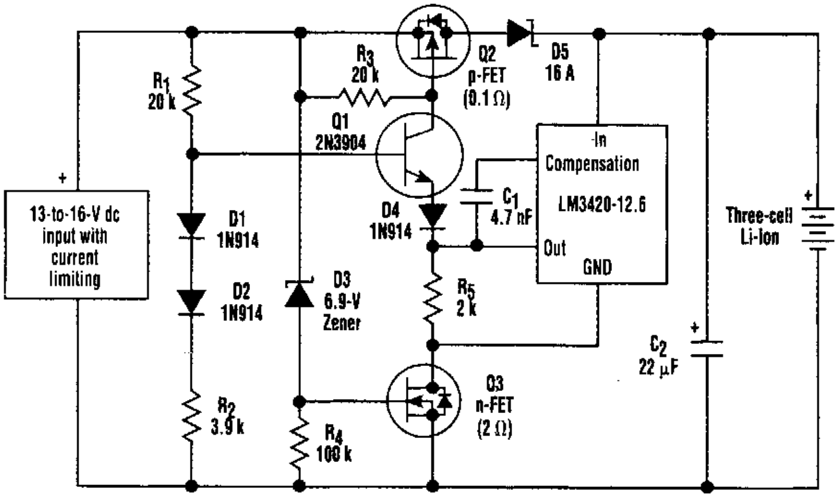


POPULAR ELECTRONICS

Fig. 13-5

This circuit uses an LM317 as a current regulator to limit charging current to a lead-acid battery. R2 should produce a 1.2-V drop at the desired limiting value of charging current.

THREE-CELL LITHIUM CHARGER

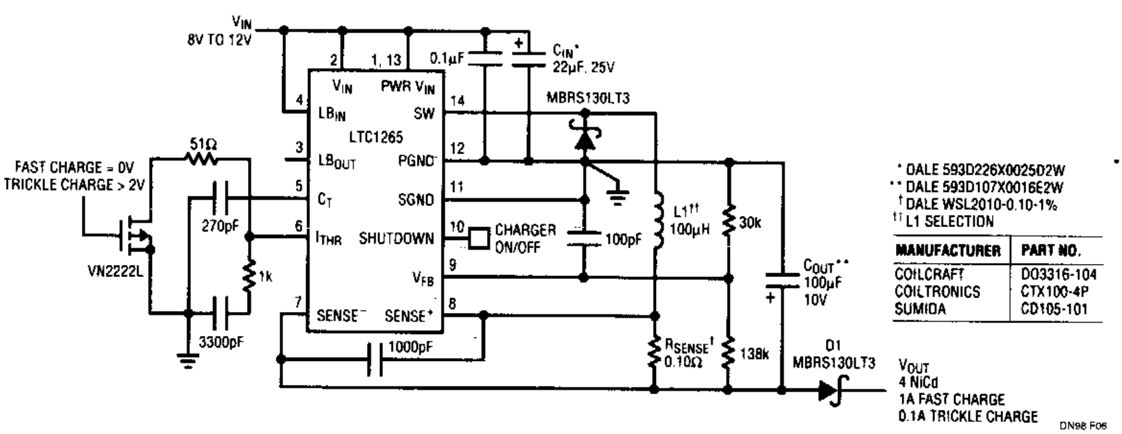


ELECTRONIC DESIGN

Fig. 13-6

This 3-A, three-cell charging circuit for lithium batteries includes a built-in on/off switch made up of Q3, R4, and D3. When a dc input is present, D3 turns on Q3, which allows current to flow through the LM3411 and Q1. If dc voltage is removed, Q3 turns off, cutting battery drain to zero.

NICAD BATTERY CHARGER

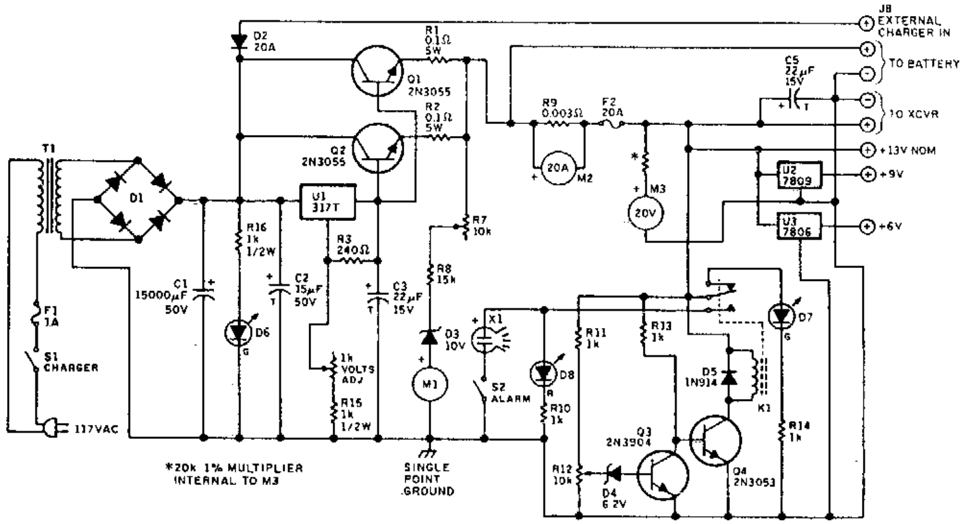


LINEAR TECHNOLOGY

Fig. 13-7

The LTC1265 is configured as a battery charger for a four-NiCad stack. It has the capability of performing a fast charge of 1 A, a trickle charge of 100 mA, or the charger can be shut off. In shut-off, diode D1 serves two purposes. First, it prevents the LTC1265 circuitry from drawing battery current and second, it eliminates "back powering" the LTC1265, which avoids a potential latch condition at power up.

BACKUP BATTERY MONITOR/CHARGER/ALARM



73 AMATEUR RADIO TODAY

Fig. 13-8

Battery Condition
Meter Calibration

Lead-Acid Battery		Lead Calcium Battery	
Color	**Voltage**	**Color**	**Voltage**
Red	11.6 and below	Red	11.6 and below
Yellow	11.6 to 12.0	Yellow	11.6 to 12.0
Green	12.0 to 13.8	Green	12.0 to 13.5
Red	13.8 and higher	Red	13.5 and higher

Charging voltage is constant at the normal full-charge level, so the charging current drops as full charge is approached, and full charge is maintained with a trickle current. The charging voltage can be adjusted between approximately 10 and 15 Vdc to accommodate lead-acid (13.8 V) or lead-calcium (13.2 V, 13.5 V maximum) deep-cycle storage batteries.

A separate connection is provided so that an external charger can be used when greater than 3 A is needed to charge a partially discharged battery. Internal circuitry will maintain the charging voltage to the battery at the nominal full-charge voltage level, regardless of the voltage supplied by the external charger, which will be 2 V or more greater than that applied by the regulator to the storage battery. Warning: do not fast-charge deep-cycle storage batteries!

A pair of meters calibrated to indicate 20 Vdc and 20 Adc full-scale monitor voltage and current when battery power is used.

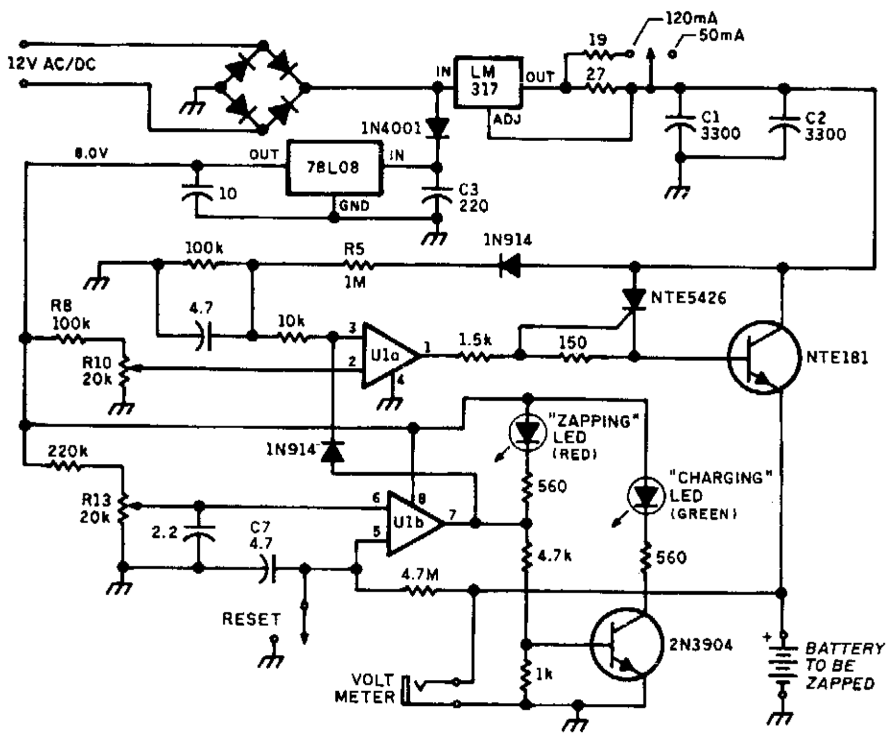
A separate, suppressed zero, expanded-scale meter calibrated over the range of about 10 to 15 Vdc allows immediate and constant indication of the state of charge of the station's backup battery. This meter scale is calibrated in bands of red, yellow, and green, as explained in the table. The narrow yellow segment is based on the assumption that solid-state transceivers might not operate properly below +12 Vdc. The internal power supply is used to calibrate this meter. A DMM should be used for greatest accuracy.

BACKUP BATTERY MONITOR/CHARGER/ALARM (*Cont.*)

An alarm circuit is included to indicate when the battery has been discharged by 60 percent to the 11.6-Vdc level. When battery voltage is above 11.6 V, the green LED will be illuminated; when voltage falls to 11.6 V, the green LED goes out and the red LED lights. A piezo audible alarm sounds at this low-voltage level unless silenced by the toggle switch controlling it.

A pair of fixed three-terminal regulators are included to provide +9 and +6 Vdc.

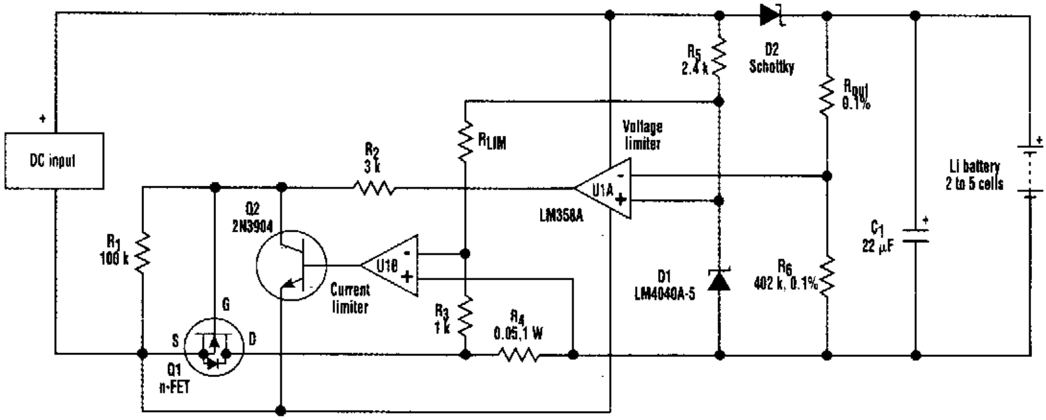
NICAD CHARGER/ZAPPER



73 AMATEUR RADIO TODAY ***Fig. 13-9***

The NiCad charger/zapper has a built-in charger and zapper circuit to clear shorted NiCads. This circuit delivers a high-current pulse to trim out internal shorts.

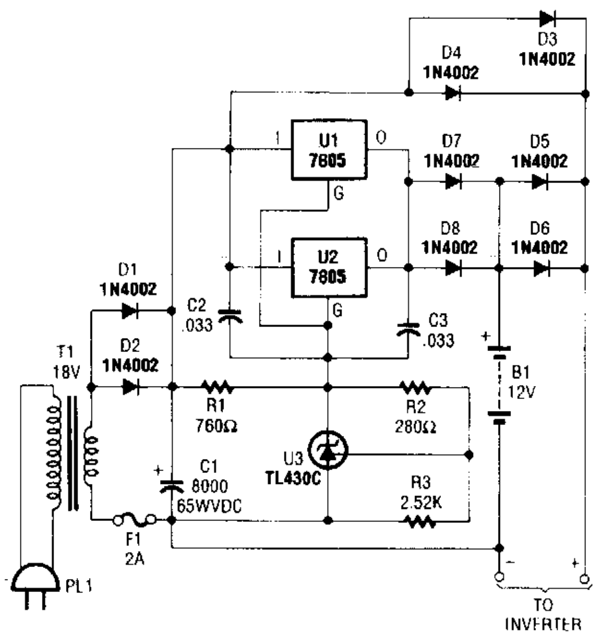
2- TO 5-CELL LITHIUM BATTERY CHARGER



ELECTRONIC DESIGN *Fig. 13-10*

A more generally applicable circuit-design concept for recharging lithium batteries could easily accommodate different cell types and various numbers of cells. That's because both the charger output-voltage set point and current limit, or maximum charging current, can be adjusted by simply changing a resistor.

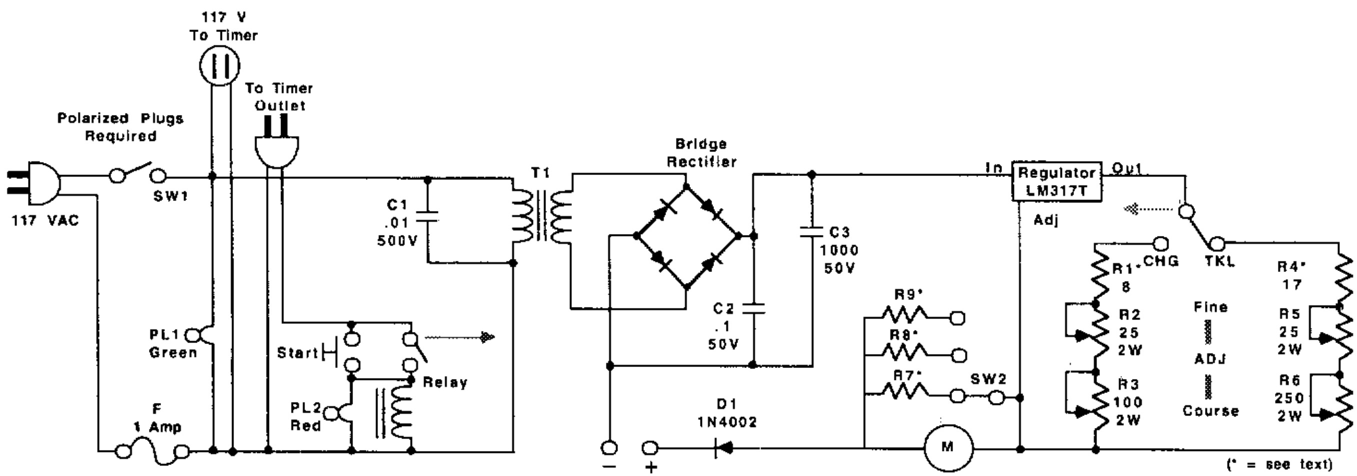
LEAD-ACID TRICKLE CHARGER



POPULAR ELECTRONICS *Fig. 13-11*

This lead-acid battery trickle charger can be used as a stand-alone circuit (for alarm systems and such) or combined with the circuit in the figure to create an emergency lighting system.

NICAD BATTERY CHARGER



73 AMATEUR RADIO TODAY

Fig. 13-12

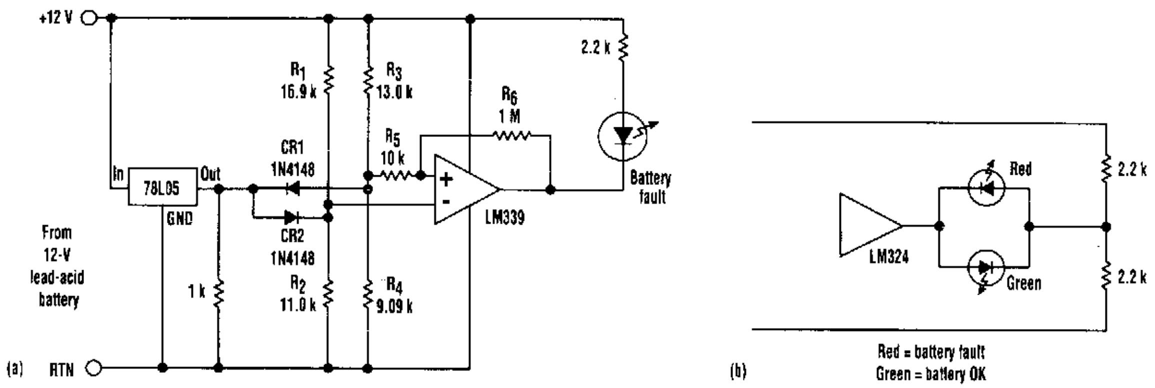
This circuit has a current regulator and uses an external timer to control the charging rate.

14

Battery Monitor Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

BATTERY MONITOR



ELECTRONIC DESIGN

Fig. 14-1

One typical application for the detector involves monitoring a lead-acid battery. It indicates a fault when the battery voltage is outside an 11- to 14-V window. Because the circuit is powered by the battery, the input and reference were switched to keep the comparator inputs within its common-mode range.

The circuit's reference is 5.0 V. The resistor values in divider, R1/R2 were selected to produce 5.5 V at the inverting input when the battery voltage is 14.0 V. Divider R3/R4 is set to produce 4.5 V at the noninverting input when the battery voltage is equal to 11.0 V.

When the battery voltage is within the window, the noninverting input is more positive than the inverting input which is clamped at 4.5 V by CR2, the noninverting input continues below that, the comparator's output goes low, and the LED turns on. When the battery voltage rises above 14 V, the noninverting input is clamped at 5.5 V by CR1, the inverting input continues above that, the comparator output again goes low, and the LED turns on. Resistors R5 and R6 show that hysteresis might be added to this circuit in a conventional manner.

If an op amp, such as an LM324 is used as the comparator, two LEDs can be implemented. The green LED will turn on when the battery voltage is within the window, and the red LED turns on when the battery voltage is outside the window.

BATTERY BUTLER

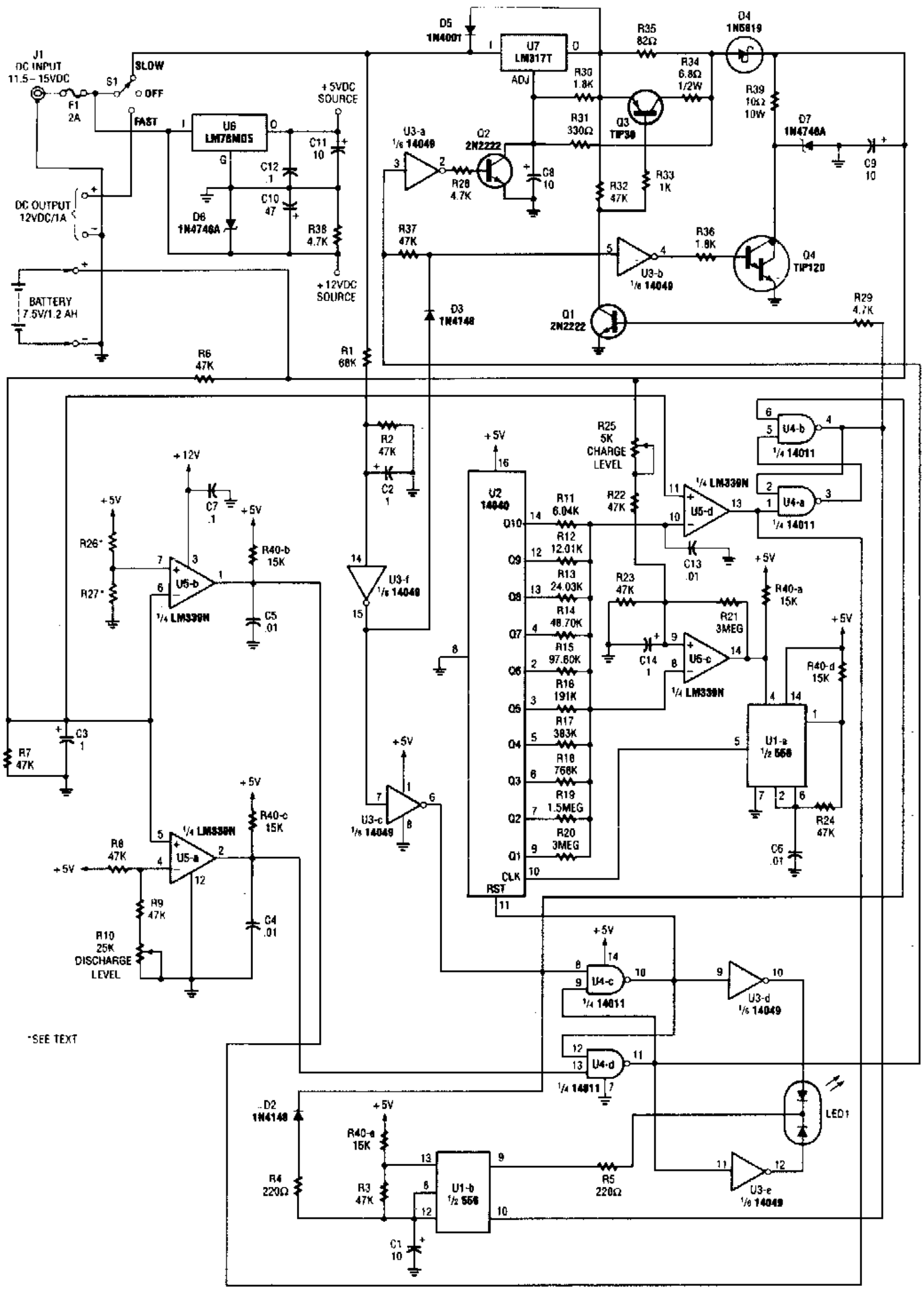
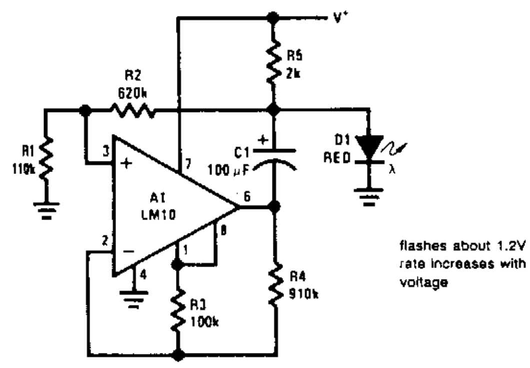


Fig. 14-2

The battery butler solves the common problems associated with the maintenance and operation of NiCad batteries. The battery butler, by initially discharging a NiCad battery to a preset point, reduces the possibility of the "memory" effect occurring. Once discharged, a battery is then usually charged at 25% and reduce the internal cell pressure increase by 40% or more. Once the battery is fully charged, a trickle charge is provided to maintain the battery in a fully charged state. The battery butler circuit can be bypassed, and the existing fast-charger used, if needed.

UNDERVOLTAGE INDICATOR FOR SINGLE CELL



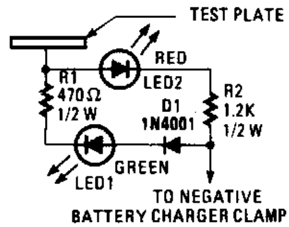
NATIONAL SEMICONDUCTOR *Fig. 14-3*

When operating with a single cell, it is necessary to incorporate switching circuitry to develop sufficient voltage to drive the LED. A circuit that accomplishes this is drawn in the figure shown. Basically, it is a voltage-controlled asymmetrical multivibrator with a minimum operating threshold given by:

$$V_{TH} = \frac{R_4\,(R_1 + R_2)}{R_1\,(R_3 + R_4)}\,V_{REF}$$

Above this threshold, the flash frequency increases with voltage. This is a far more noticeable indication of a deteriorating battery than merely dimming the LED. In addition, the indicator can be made visible with considerably less power drain. With the values shown, the flash rate is 1.4 sec −1 at 1.2 V with 300-μA drain and 5.5 sec −1 at 1.55 V with 800-μA drain. Equivalent visibility for continuous operation would require more than 5-mA drain.

BATTERY CHARGER PROBE



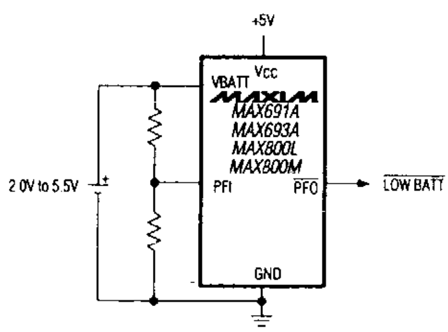
POPULAR ELECTRONICS *Fig. 14-4*

This battery-charger probe can keep you from damaging batteries or yourself by testing to see if the charger is already on and/or connected improperly.

To use the probe, the positive cable clamp is first connected to the positive battery terminal. Then, the test plate is touched to the negative terminal of the battery. If the battery is connected properly, current will pass from the test plate through R1, LED1, D1, the negative charger, and into the positive side of the batteries. If LED1 (the green LED) lights, you can clamp on the negative lead and turn on the charger.

If the terminals are reversed, current will flow in the opposite direction, causing LED2 to light, warning you of danger. When the cable is reversed, D1 protects LED1 from excessive reverse voltage. If that happens, immediately turn the power off, and right the cable connections. Finally, if the battery charger is on, both LEDs will light because chargers actually produce pulsating dc and rely on the battery to act as a filter.

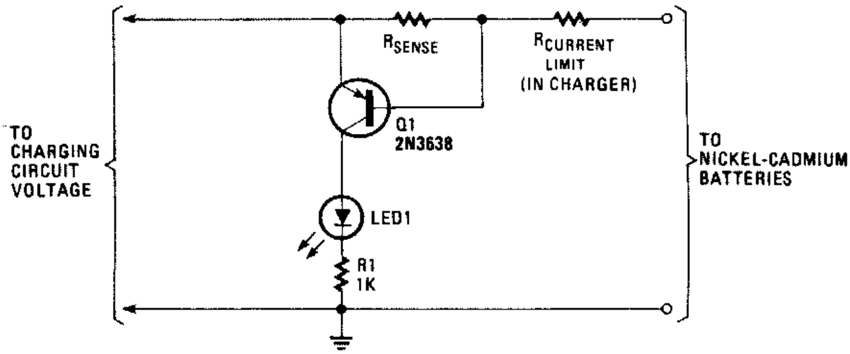
LOW-BATTERY CIRCUIT



MAXIM *Fig. 14-5*

A Maxim MAX691A series IC allows low-battery detection.

BATTERY CHARGE INDICATOR



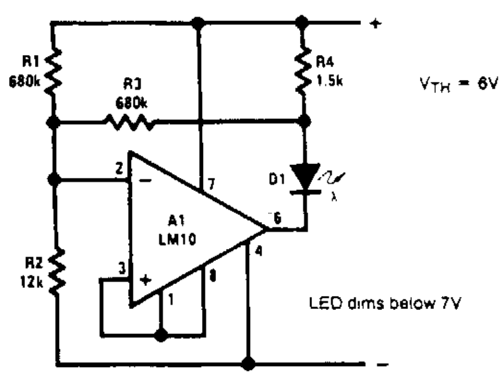
RADIO-ELECTRONICS

Fig. 14-6

When a battery is charging, a voltage drop across R_{SENSE} causes Q1 to conduct, and lights LED1. R_{SENSE} should be chosen as follows:

$$R_{SENSE}\text{ (ohms)} = \frac{0.65}{I_{CHARGE}\text{ (amps)}}$$

BATTERY STATUS INDICATOR



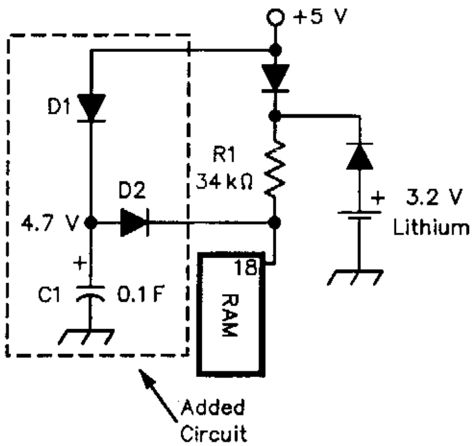
NATIONAL SEMICONDUCTOR

Fig. 14-7

In battery-powered circuitry, there are some advantages to having an indicator to show when the battery voltage is high enough for proper circuit operation. This is especially true for instruments that can produce erroneous data.

The battery status indicator is designed for a 9-V source. It begins dimming noticeably below 7 V and it extinguishes at 6 V. If the warning of incipient battery failure is not desired, R3 can be removed and the value of R_1 is halved.

LITHIUM MEMORY BACKUP BATTERY REPLACEMENT

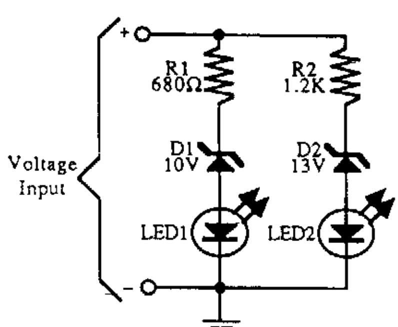


QST

Fig. 14-8

Physically very small high-capacitance capacitors are available for memory backup. Here, a 0.1-F (100,000 μF) capacitor and two diodes replace the lithium battery. The lithium battery can be retained as well, providing double backup.

BATTERY-CONDITION INDICATOR FOR 12-V BATTERIES



McGRAW-HILL

Fig. 14-9

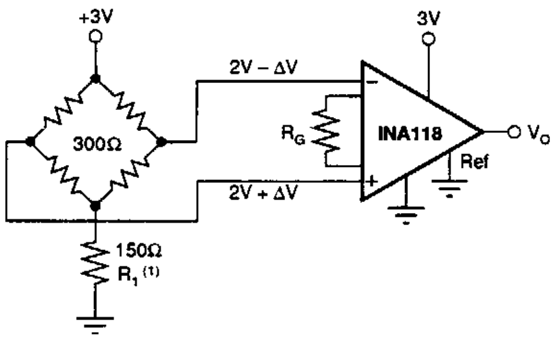
A simple battery condition indicator. Choose the Zener diodes to provide a "window" for over/under voltage indication.

15

Bridge Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SINGLE-SUPPLY BRIDGE AMPLIFIER



NOTE: (1) R_1 required to create proper common-mode voltage, only for low voltage operation — see text.

DESIRED GAIN	R_G (Ω)	NEAREST 1% R_G (Ω)
1	NC	NC
2	50.00k	49.9k
5	12.50k	12.4k
10	5.556k	5.62k
20	2.632k	2.61k
50	1.02k	1.02k
100	505.1	511
200	251.3	249
500	100.2	100
1000	50.05	49.9
2000	25.01	24.9
5000	10.00	10
10000	5.001	4.99

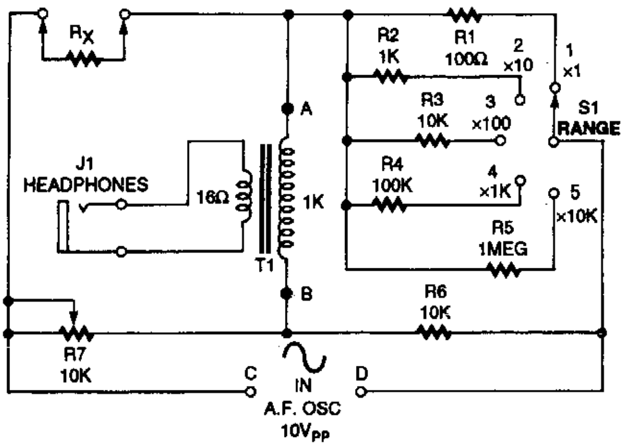
BURR-BROWN *Fig. 15-1*

The INA118 can be used on single-power supplies of +2.7 to +36 V. The figure shown is a basic single-supply circuit. The output Ref terminal is connected to ground. Zero differential input voltage will demand an output voltage of 0 V (ground). Actual output voltage swing is limited to approximately 35 mV above ground, when the load is referred to ground as shown. The typical performance curve "Output Voltage vs. Output Current" shows how the output voltage swing varies with output current.

With single-supply operation, $+V_{IN}$ and $-V_{IN}$ must both be 1.1 V above ground for linear operation. You cannot, for instance, connect the inverting input to ground and measure a voltage connected to the noninverting input.

To illustrate the issues affecting low-voltage operation, consider the circuit in the figure. It shows the INA118, operating from a single 3-V supply. A resistor in series with the low side of the bridge ensures that the bridge output voltage is within the common-mode range of the amplifier's inputs.

WHEATSTONE BRIDGE



POPULAR ELECTRONICS

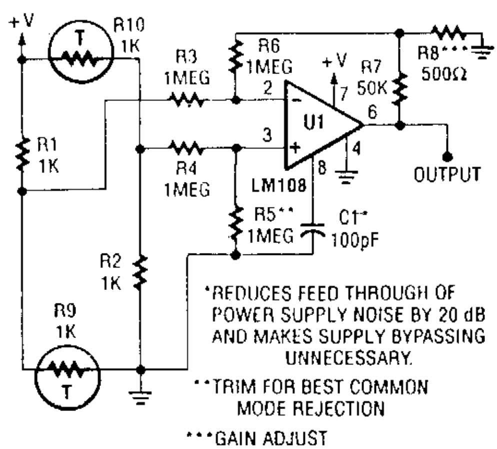
Fig. 15-2

This circuit can be used to measure resistances. R7 is calibrated and fitted with an indicator dial, then:

$$\frac{R_x}{(R_1 \text{ through } R_5)} = \frac{R_7}{R_6} \text{ or } R_x = \frac{R_7}{R_6} \times (R_1 \text{ through } R_5)$$

A frequency of 1 kHz for the audio oscillator is usually used.

BRIDGE AMPLIFIER WITH LOW NOISE COMPENSATION



POPULAR ELECTRONICS

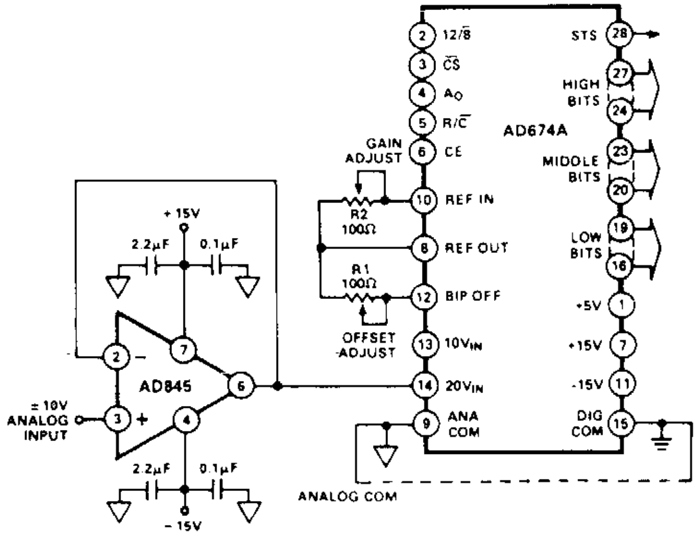
Fig. 15-3

16

Buffer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

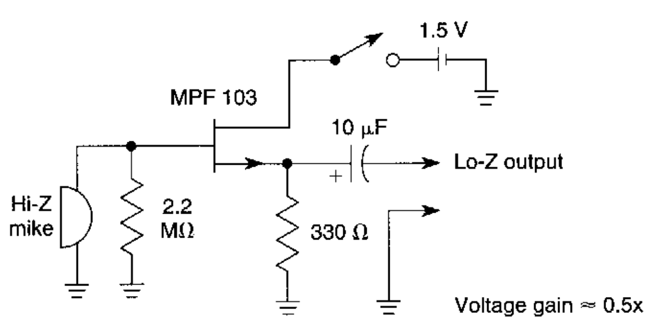
UNITY-GAIN ADC BUFFER



ANALOG DEVICES **Fig. 16-1**

This buffer is suitable for ADCs of 12 bits with conversion times of 5 μs or greater. The wide bandwidth of the AD845 ensures a low output impedance at higher frequencies in the voltage follower (buffer) configuration.

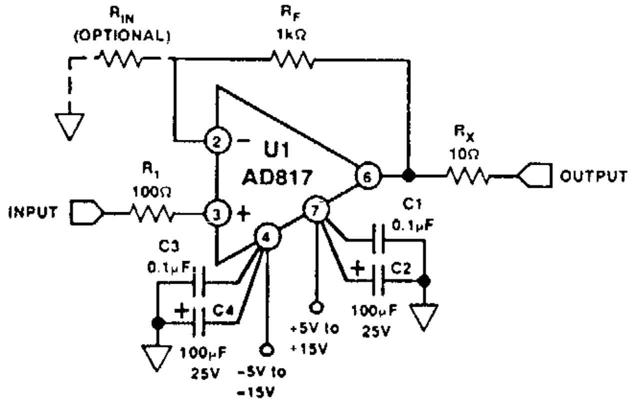
HI-*Z* MICROPHONE BUFFER AMPLIFIER



WILLIAM SHEETS **Fig. 16-2**

A low impedance output from a high-*Z* microphone can be obtained with this circuit. No voltage gain is obtained, but a power gain is obtained because the output impedance is much lower (300 Ω), with –6-dB voltage gain.

WIDEBAND GENERAL-PURPOSE BUFFER

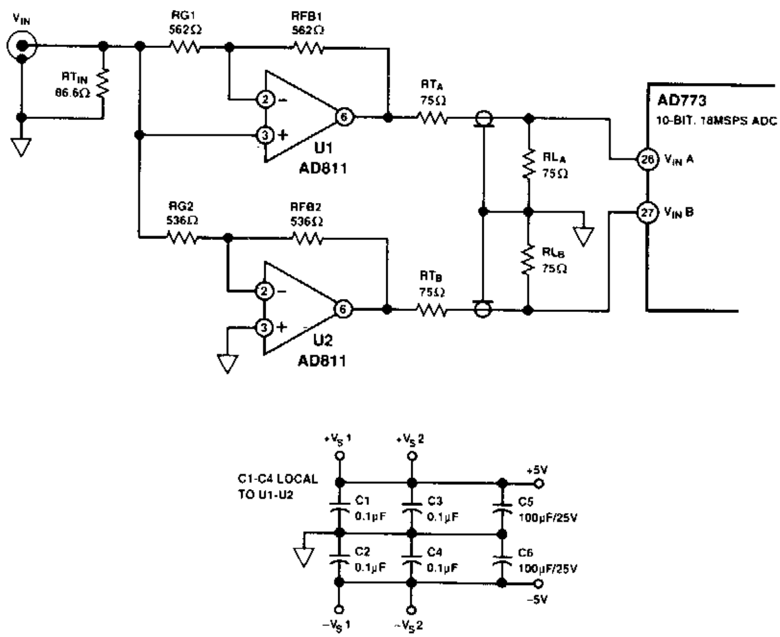


ANALOG DEVICES

Fig. 16-3

This circuit has unity gain and response up to 70 MHz. U1 is an Analog Devices AD817.

ADC BUFFER

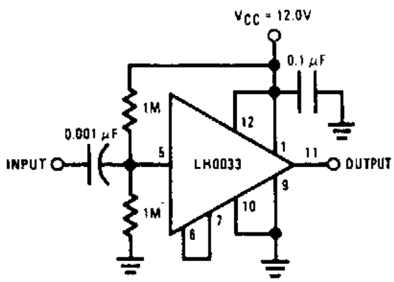


ANALOG DEVICES

Fig. 16-4

Useful for driving high-speed, 10-bit ADCs, this circuit was developed to drive an 18-MSPS 10-bit ADS. It works from ±5-V supplies.

SINGLE-SUPPLY ac BUFFER AMPLIFIER

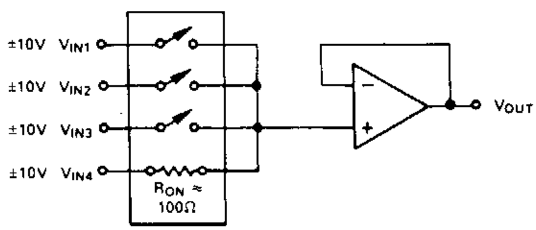


This buffer might be used with a single supply without special considerations. The input is dc biased to mid-operating point and is ac coupled. Its input impedance is approximately 500 kΩ at low frequencies. Note that for dc loads referenced to ground, this quiescent current is increased by the load current set at the input dc bias voltage.

NATIONAL SEMICONDUCTOR

Fig. 16-5

ANALOG NONINVERTING SWITCHED BUFFER

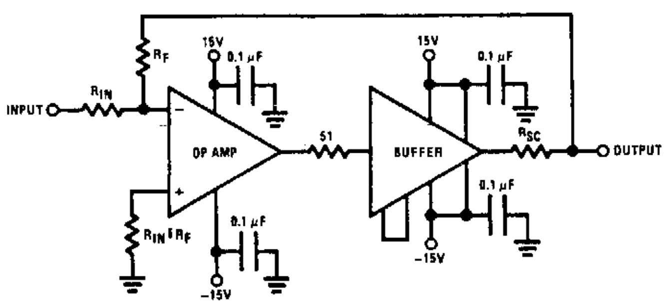


Here is noninverting solution.

ANALOG DEVICES

Fig. 16-6

VOLTAGE FOLLOWER

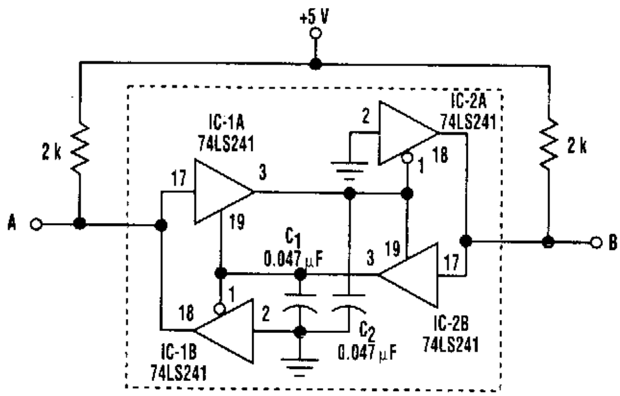


Op Amp	Recommended Buffer	
LM101, LM108, LM741, LF151 LH0022, LH0042, LH0052	LH0002	
LF155, LF156, LF157, LH0024, LH0032	LH0033	$R_{SC} \frac{V_S}{I_{SC}}$
LH0024, LH0032	LH0063	

NATIONAL SEMICONDUCTOR

Fig. 16-7

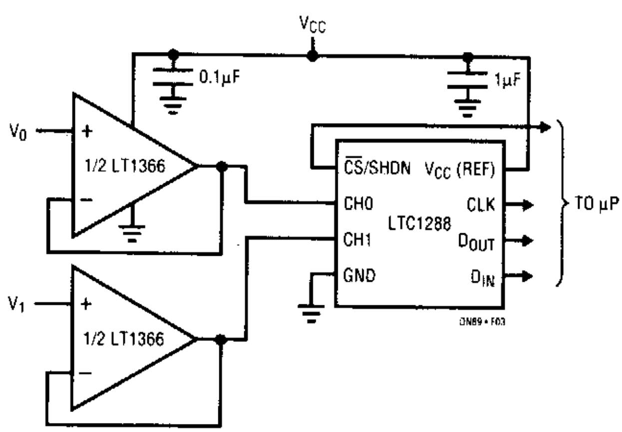
SIMPLE BIDIRECTIONAL BUFFER DESIGN



ELECTRONIC DESIGN *Fig. 16-8*

This circuit shown in the figure uses two 74LS241s. When both input/output lines are high, IC-1A and IC-2B turn on, and C1 and C2 are charged to high voltage. Meanwhile, IC-1B and IC-2A are off to prevent a logic "1" latch.

BUFFER FOR A/D CONVERTERS



LINEAR TECHNOLOGY *Fig. 16-9*

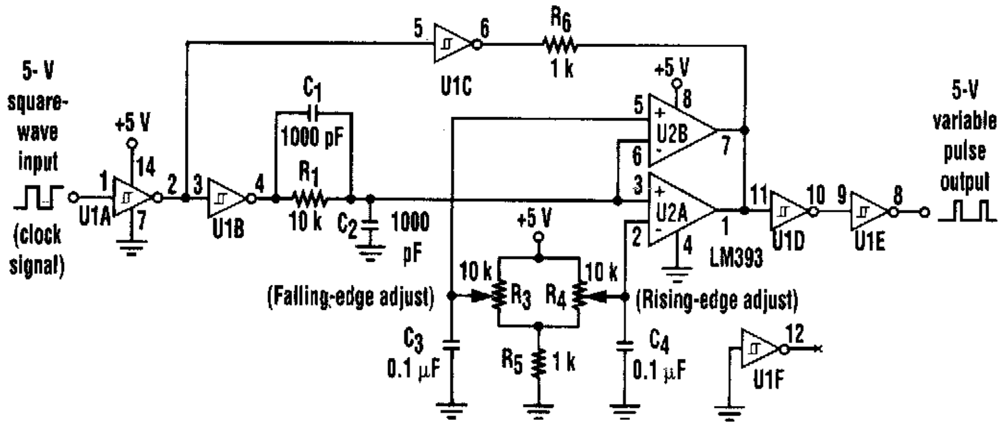
This circuit uses an LT1366 driving an LTC1288 two-channel micropower A/D. The LTC1288 can accommodate voltage references and input signals equal to the supply rails. The sampling nature of this A/D eliminates the need for an external sample-and-hold, but might call for a drive amplifier because of the A/D's 12-μs settling requirement. The LT1366's rail-to-rail operation and low-input offset voltage make it well suited for low-power, low-frequency A/D applications. In addition, the op-amp's output settles to 1% in response to a 3-mA load step through 100 pF in less than 1.5 μs.

17

Clock Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SET TIME WINDOWS WITHIN A CLOCK



ELECTRONIC DESIGN

Fig. 17-1

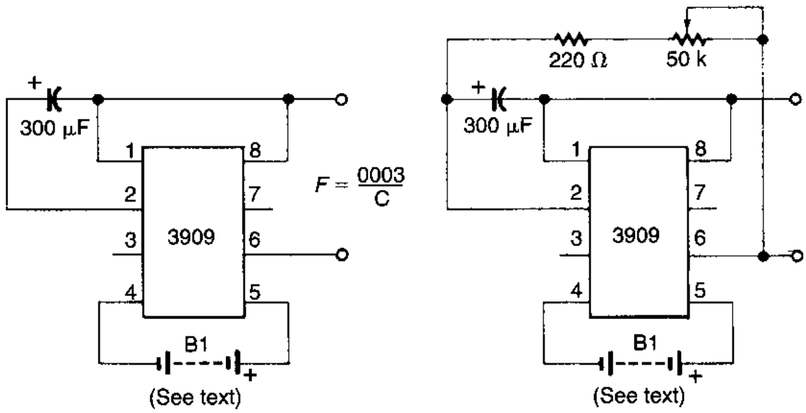
At times, it is necessary to produce pulses of adjustable width and whose start times might vary with reference to a master clock. The input signal is inverted and buffered by U1A, while U1B and U1C reinvert the signal to produce square, buffered renditions of the input signal. Potentiometers R3 and R4 set references for the comparators.

The input polarity of U2A keeps its output transistor turned on until the voltage at the noninverting input exceeds the reference set by R4 (the rising edge adjustment). When this reference voltage is exceeded, the output transistor is turned off and the output signal is pulled up via R6. Meanwhile, the input polarity of U2B keeps its output transistor turned off until the voltage at the inverting input exceeds the reference set by R3 (the falling edge adjustment). When this reference voltage is surpassed, the output transistor of U2B is turned on, pulling the output signal low through the wired-OR configuration of U2. The output of U2 is then double-inverted and buffered by U1D and U1E. What results is a pulse whose start time (rising edge) can be adjusted by R4, and whose stop time (falling edge) can be adjusted by R3.

The output of the comparators is pulled up to the input waveform through resistor R6 to U1C. This prevents the comparators from switching during the low cycle of the input waveform, regardless of the positions of R3 and R4. This has the effect of "locking out" changes during the low period of the input signal, and would probably require additional logic if it were done strictly in the digital domain.

The circuit, with the component values shown, works well between about 50 and 150 kHz.

LOW-FREQUENCY CLOCK



ELECTRONICS NOW

Fig. 17-2

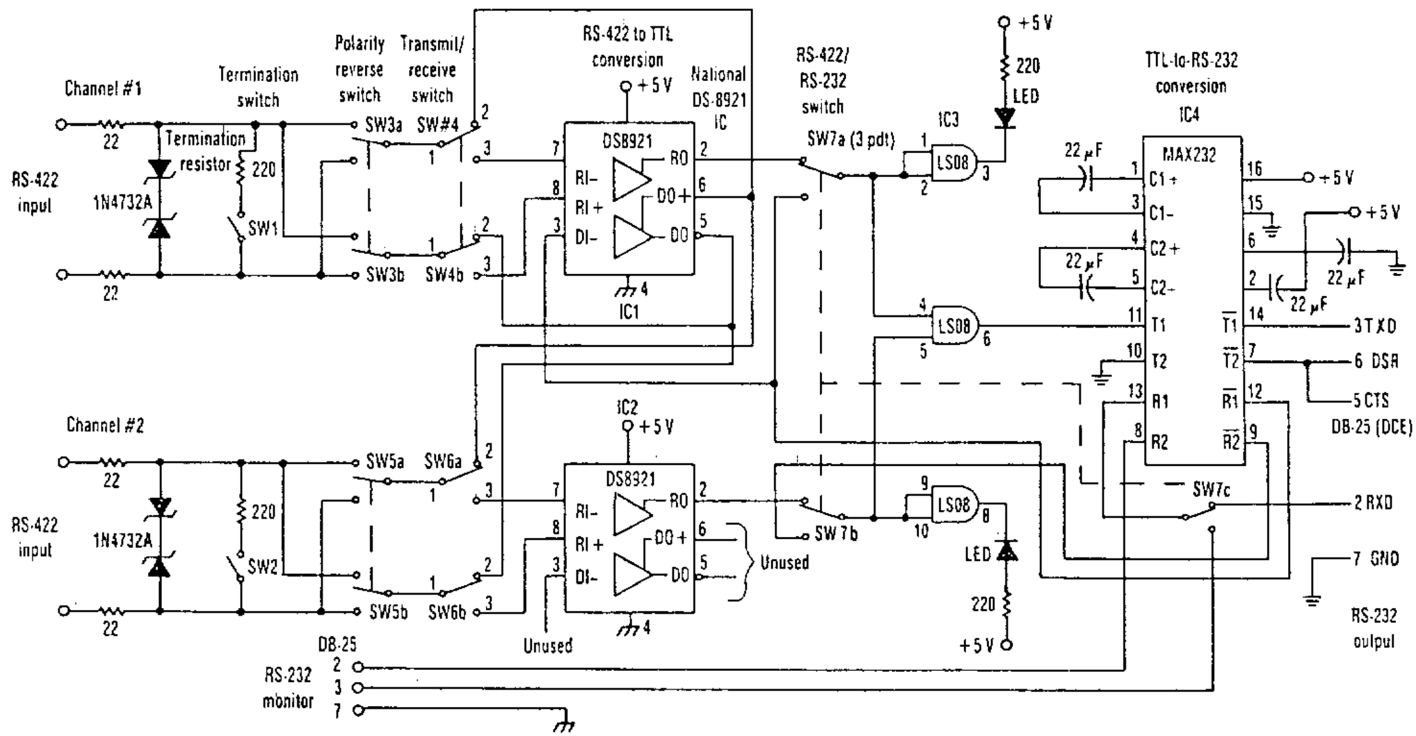
The LM3090 is an LED flasher IC that is designed to oscillate at low frequencies. The clock output of the first circuit can be changed by changing the value of the capacitor, and the second circuit lets you adjust the frequency with the trimmer. B1 can be one or two alkaline 1.5-V cells. The LM3909 can supply up to 45-mA pulses at greater than 2 V.

18

Computer-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

RS-422 TO RS-232 CONVERTER



ELECTRONIC DESIGN

Fig. 18-1

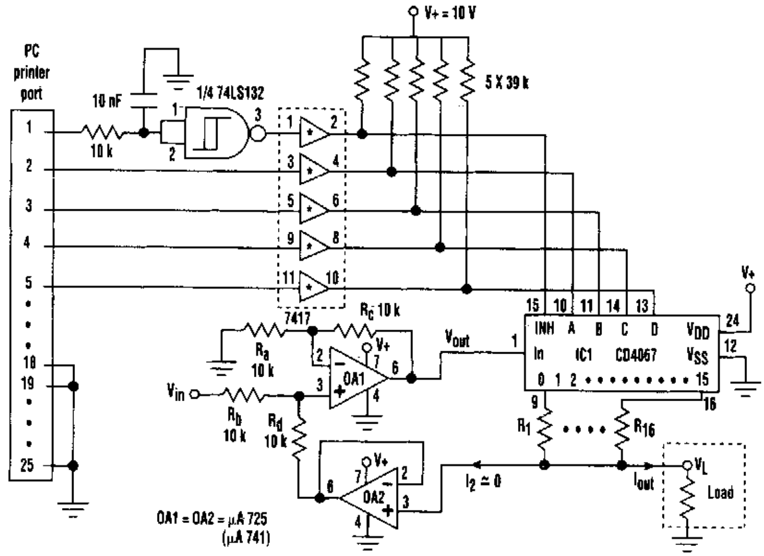
The circuit supplies two LEDs for visual indication of line activity and terminating resistors when needed. The 220-Ω resistors and 5-V zeners at the RS-422 line inputs supply circuit protection.

Switch SW7 allows the circuit monitor both transmitted and received signals when tee-connected into an RS-232 line. One function of this optional feature is the ability to test a software-locking device that connects to the COM1 port on an IBM PC.

PRINTER PORT

REGISTOR SELECTION CODES

Data	Selected resistor
0X00	R_1
0X01	R_2
0X02	R_3
0X03	R_4
0X04	R_5
0X05	R_6
0X06	R_7
0X07	R_8
0X08	R_9
0X09	R_{10}
0X0A	R_{11}
0X0B	R_{12}
0X0C	R_{13}
0X0D	R_{14}
0X0E	R_{15}
0X0F	R_{16}



ELECTRONIC DESIGN

Fig. 18-2

A 16-step programmable current generator can be modified so that it's controllable by a printer port. This is done by switching the resistor connected between the output of the generator's OA1 op amp and the input of OA2. The CMOS single 16-channel analog multiplexer (IC1) chooses one resistor at a time, in accordance with the code sent by the printer port through four of its eight data-output lines (pins 2 to 9). In addition, one control line (pin 1) is used to enable the operation. As a result, 16 outputs can be selected by a 4-bit word (the table shows the relationship between data word and selected resistor).

The following must be fulfilled in order for the circuit to work as a true current generator:

$$R_2 \times R_d - R_b \times R_c = 0$$

The smaller the resistors' tolerance (especially R1 through R16), the greater the output resistance of the generator.

Because the OA2 is connected as a repeater, the current $I_2 = 0$, and only the load current flows through one of the R1 through R16 resistors. Therefore:

$$I_{out} = \frac{V_{out}}{(R_x + R_{on})}$$

where $V_{out} = V_{in}$; $X = 1...16$; and $R_{on} \leq 150\ \Omega$ (for $V_{DD} = 10$ V) is the resistance of one analog switch (CD4067) in conduction.

Therefore, the values of resistors R_1 through R_{16} can be inferred from the needed currents:

$$R_x = \left(\frac{V_{in}}{I_{out}}\right) - R_{on}$$

The Turbo C++ program also controls the current through the load.

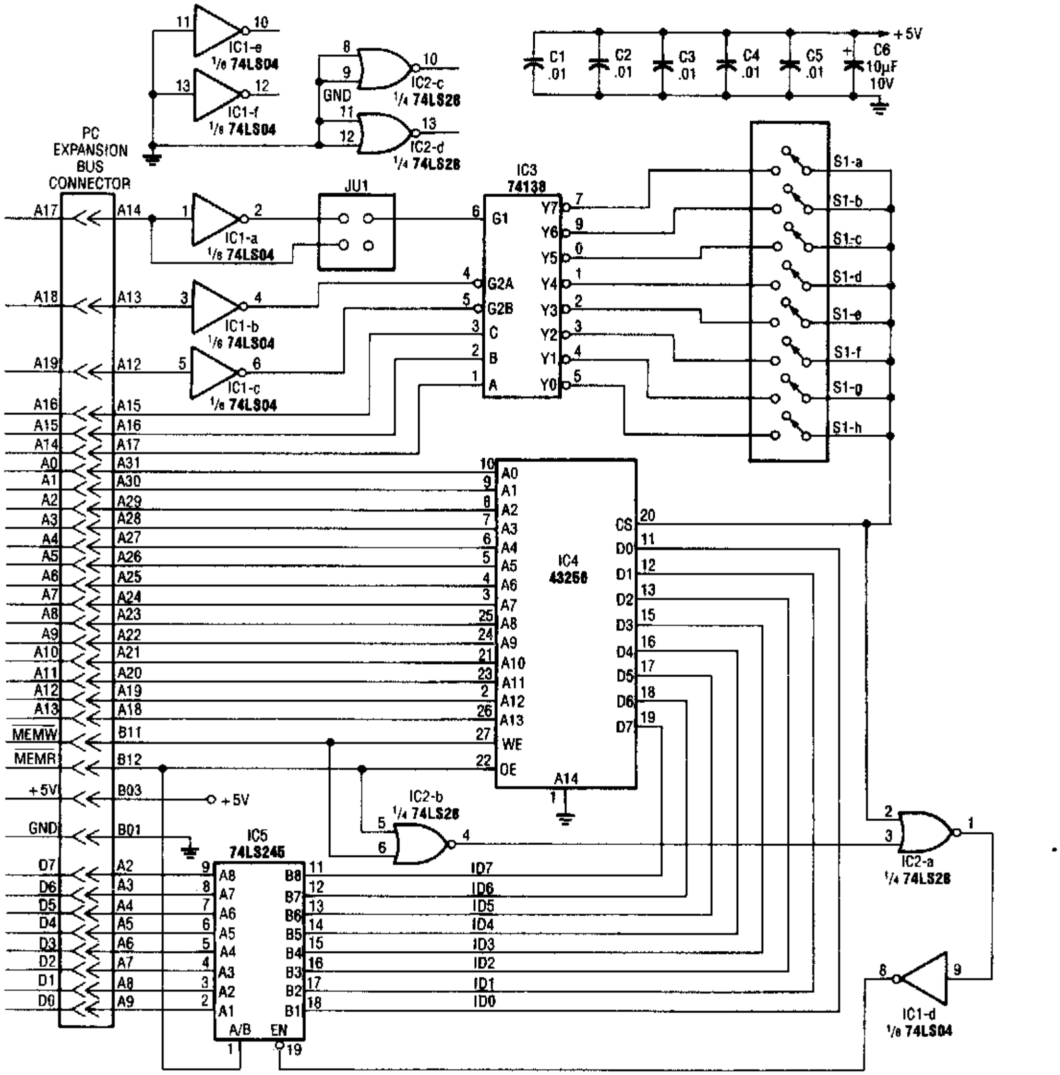
TURBO C++ CONTROL PROGRAM

```
#include <stdio.h>
#include <dos.h>

#define   OUT_PORT   0x378    /*printer output port address   */
#define   CTRL_PORT  0x37A    /*printer control port address   */

int main(void)
{
  int data;
  outport(CTRL_PORT , 0x01);    // enable operation
  delay(1);
  outportb(OUT_PORT) , data); // one of R1 -R16 selected (table)
  printf("\n\aR%d selected.",data+1);
  return 0;
}
```

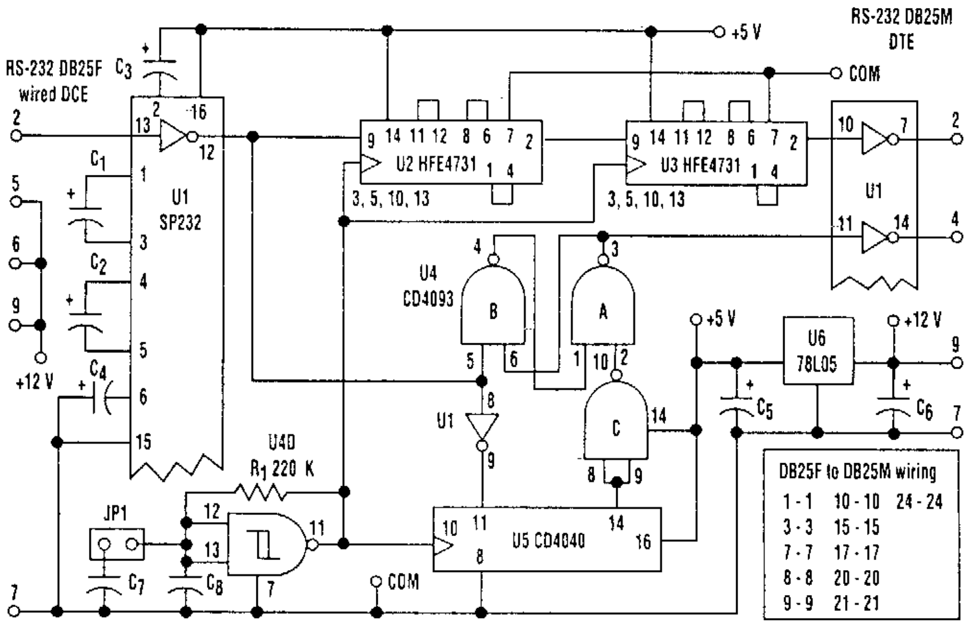
PC PASSWORD PROTECTOR



ELECTRONIC EXPERIMENTERS HANDBOOK *Fig. 18-3*

IC4, a static RAM, is mounted in a "smart" built-in switch over circuitry. This retains SRAM contents when power is off. The rest of the circuitry consists of address decoding logic and jumper JU1, used to decode a 16K address space for the 32K static RAM. Software is necessary and this is contained in the original article (see reference).

KEY WIRELESS RTS WITH DATA

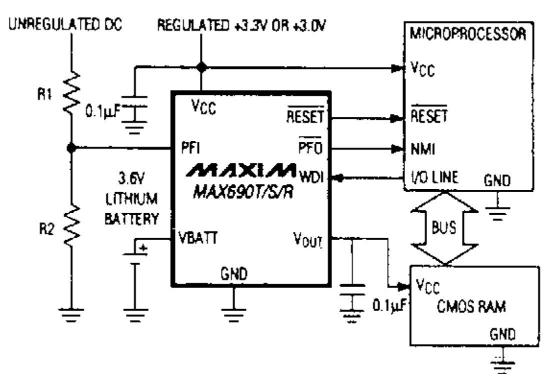


Notes: C_1-C_6 = 1μF, 35-V Ta C_7-C_8 = 100-pF disc R_1 = 5%, 1/8 W U1 = Sipex SP232 U2-U3 = Philips HFE4731

ELECTRONIC DESIGN *Fig. 18-4*

This simple keyer supplies both the RTS control and data delay needed to interface a digital radio with an RS-232, data-only system. It supports speeds to 19.2 kbits/s sync or async.

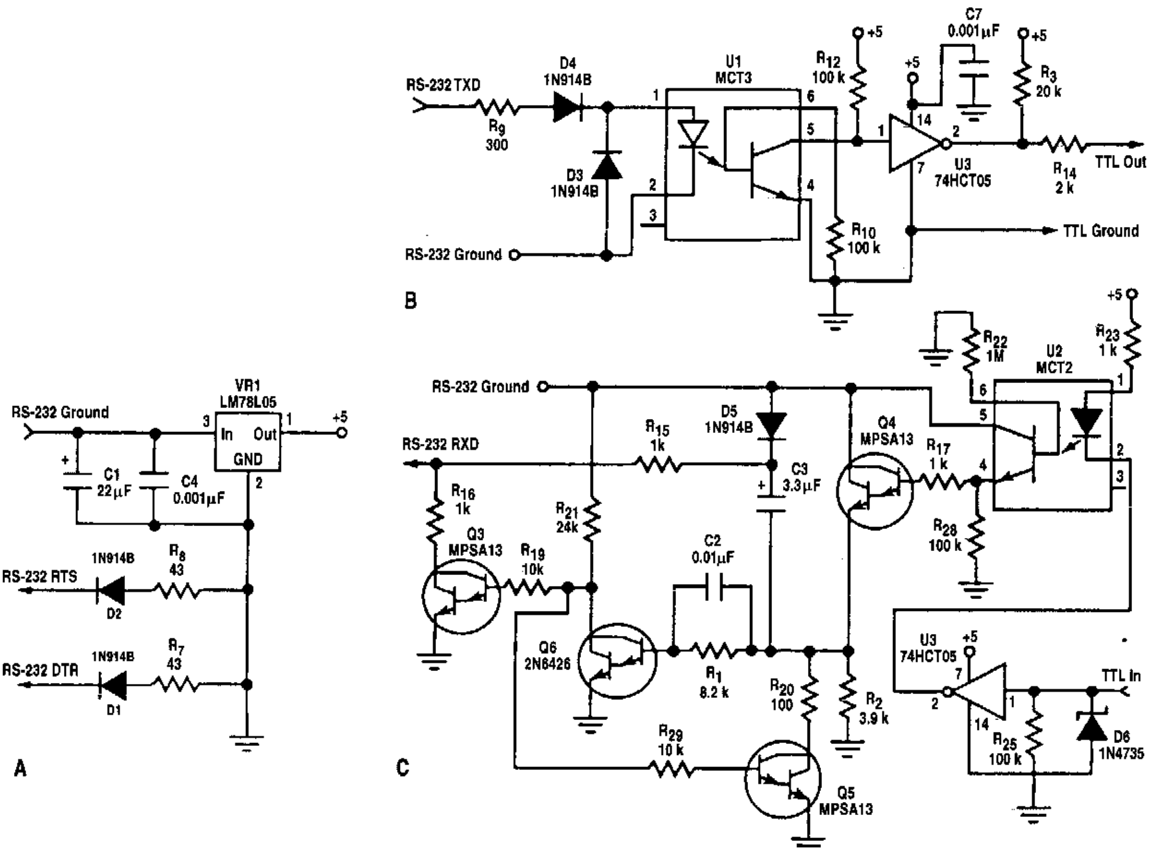
MICROPROCESSOR SUPERVISORY CIRCUIT



MAXIM *Fig. 18-5*

This circuit provides a reset pulse during power up, down, or low-voltage battery back up, switching, a watchdog timer and a 1.25-V threshold detector for power failure or power fail warning, or to monitor another supply.

COMPUTER-POWERED RS-232



ELECTRONIC DESIGN

Fig. 18-6

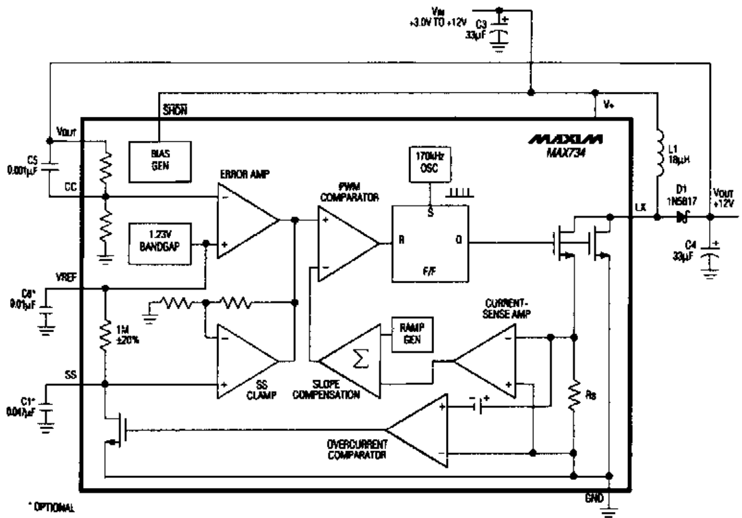
Commercializing battery-operated equipment that must interface to a computer via the RS-232 port runs into the problem of power consumption. To load the system batteries strictly to power the interface is unacceptable. An alternative is to let the computer that the device is connected to provide the interface's power. One snag is that the RS-232 specification doesn't have a power tap on the connector, but it does provide RTS and DTR (request to send and data terminal ready) signals that assert a negative voltage in their quiescent state.

Figure 18-6A shows a simple scheme of deriving a 5-V potential from the RTS and DTR signals. R7 and R8 and diodes D1 and D2 mix the return current to the RS-232 port so that the RTS and DTR drivers split the current drawn by the interface. This scheme, even from a laptop computer, can supply 12 mA to the interface. The only drawback is that the TTL device must be isolated from the computer's ground (earth ground) because the interface treats the RS-232 ground as a positive voltage.

A modified optocoupler system shifts the RS-232 level to TTL voltages (Fig. 18-6B). It will support up to 9600 bits/s. C3 is charged up to about 1 V less than the RTS voltage while the TTL line asserts a marking state. As the capacitor is charging, Q3 is biased into saturation, thus providing a negative voltage (with respect to RS-232 ground) to the RS-232 RXD line. When a spacing bit is driven from the TTL line, Q3 switches off and Q4 switches on. This biases Q4's emitter up to the RS-232 ground. That ground potential is summed with C3's charge to create an RS-232-compatible spacing signal (approximately 1 V less than a $-V_{RTS}$).

The discharge rate on C3 is limited by R15 to prevent the signal sag from becoming a problem down to 110 bits/s. The C2/R1 time constant must be fairly close (within 4 times) to the C3/R15 time constant to ensure that Q3 turns off correctly.

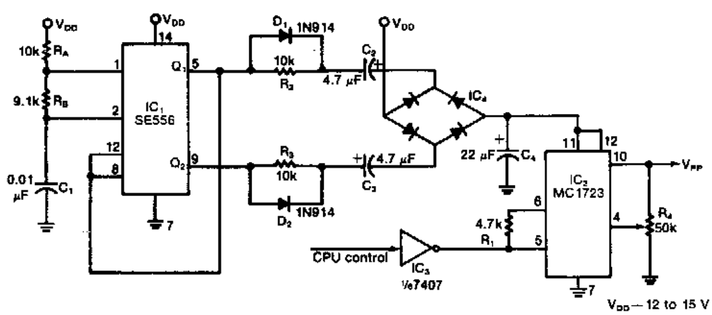
+12-V FLASH MEMORY PROGRAMMING SUPPLY



MAXIM *Fig. 18-7*

The MAX734 can deliver up to 120 mA @12 V from a +5-V supply, using few external components. This supply can also be used for other applications than memory programming. A logic level is used for shutdown. Efficiency is about 85%.

EEPROM PROGRAMMING DOUBLER CIRCUIT



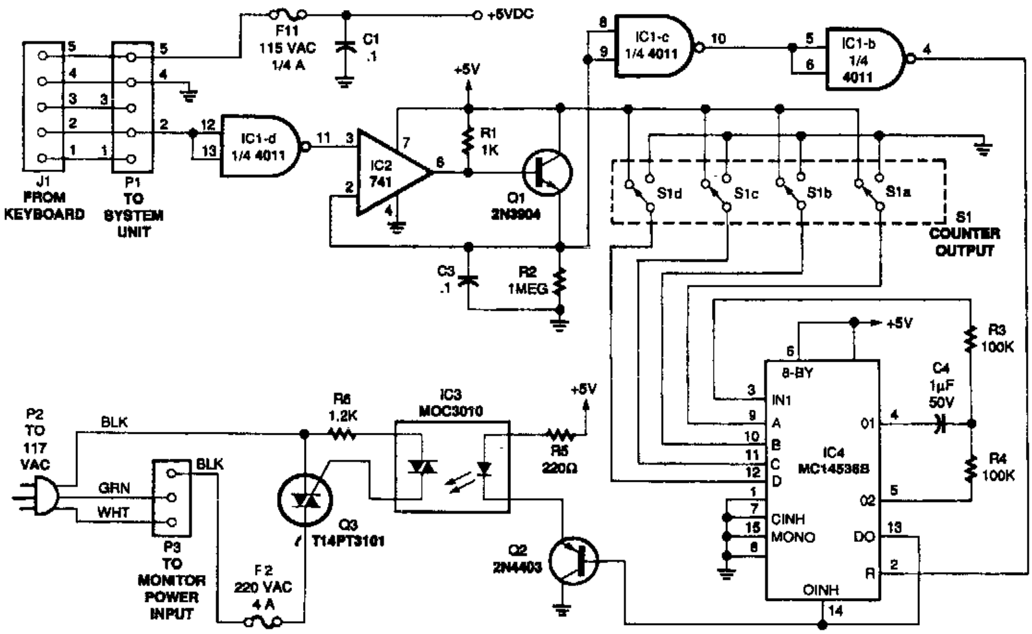
ELECTRONIC DESIGN *Fig. 18-8*

Even though electrically erasable PROMs offer the convenience of single-byte write and erase operations, the parts require a somewhat nonstandard programming voltage—21 V. A simple circuit that develops the appropriate voltage from a computer system's standard 12-to-15-Vdc supply, remedies the problem nicely. Moreover, it permits the programming voltage to be pulsed under the control of an external CPU.

As shown in the figure, the chip uses its complementary outputs, Q1 and Q2 to trigger a bridge rectifier through capacitors C2 and C3. Resistors R2 and R3 and diodes D1 and D2 limit the current and protect IC1 from spikes from C2 and C3. If required, the regulator will deliver up to 150 mA.

Circuit IC3 is an open-collector TTL gate whose output, when low, disables IC2 and causes it to put out 5 Vdc. The regulator delivers the 21-V programming pulse.

MONITOR POWER SAVER FOR COMPUTERS



ELECTRONICS NOW *Fig. 18-9*

The circuit monitors PC keyboard activity through five-pin DIN connector J1. When the user presses a key, the keyboard sends a series of negative-going pulses on pin 2. In conjunction with Q1 and C3, the op amp essentially functions as an integrator, which stretches the continually varying periods of the input pulses to a relatively constant period with a higher average dc value.

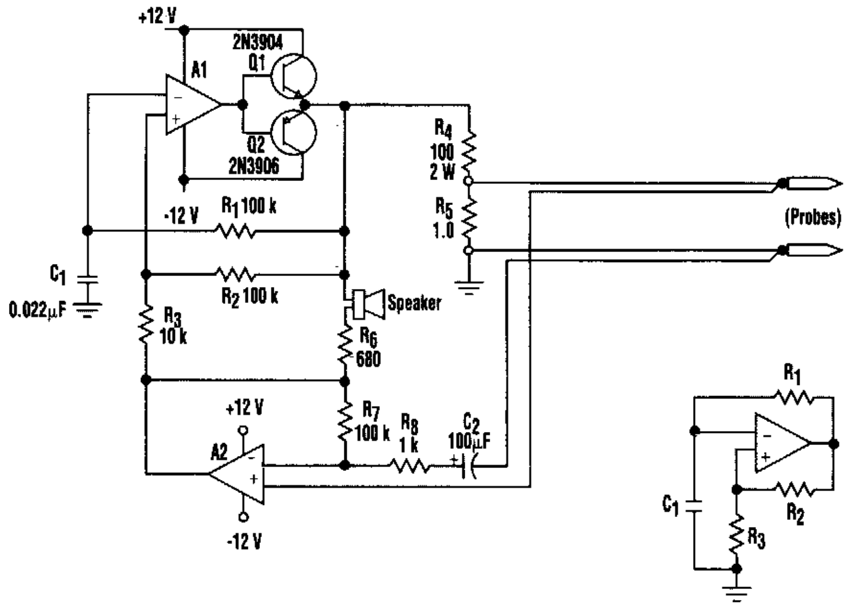
Inverters IC1-c and IC1-b buffer the peak detector's output to trigger IC4, an MC14536B programmable timer.

19

Continuity Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SHORT-CIRCUIT BEEPER



ELECTRONIC DESIGN *Fig. 19-1*

This design offers a way to trace resistance in the milliohm range, right to a short between bridged traces beneath a solder mask. It simply translates resistance into an audible tone, which increases in pitch as the measured value approaches zero.

In the classic op-amp multivibrator (shown in the inset), oscillation frequency is determined not only by the R1/C1 time constant, but also by the hysteresis set by the R_2/R_3 resistance ratio. A1 in the main figure, with current boosters Q1 and Q2, is this same configuration.

Assuming a virtual ground at the output of A2, free-run frequency is about 1 kHz—quite audible through a tiny 8-Ω speaker. Q1 and Q2 deliver a ±10-V squarewave to R4, dumping a ±100 mA through a short circuit placed across the probe tips. R5 ensures that open circuit voltage never exceeds ±0.1 V.

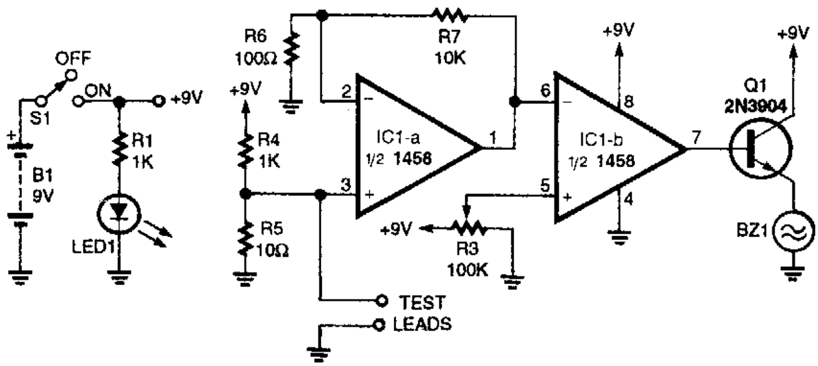
A2 monitors the voltage between the probes. The differential input must have its own separate path to the probe tips to eliminate test lead resistance from the measurement. Miniature "zip-cord" sold as loudspeaker wire makes a tidy two-conductor test lead.

When the probes are open, A2's gain equals the R4/R5 divider loss, and the output of both amplifiers is identical. This has two effects: first, hysteresis is greatly increased and the frequency falls to a low growl; second, the loudspeaker that bridges the two in-phase outputs is effectively silenced.

The dead short across the probe tips will return nothing to A2 and the circuit will squeal at its nominal 1-kHz rate. Anything less than a perfect short produces some output from A2, increasing multivibrator hysteresis and lowering the pitch. The circuit has so much "leverage," and the ear is so sensitive to pitch changes in this range, that it's easy to resolve minute resistance differences.

Any general-purpose op amp will suffice in this circuit—a couple of 741s or an equivalent dual.

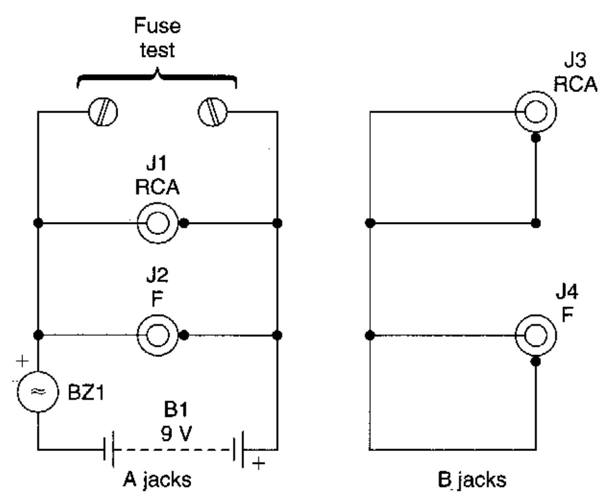
ADJUSTABLE CONTINUITY TESTER



ELECTRONICS NOW ***Fig. 19-2***

A problem with most continuity testers is that the exact decision point (circuit resistance) between continuity and open is indefinite. This circuit allows setting of this point to a known resistance between 1 and 50 Ω.

SIMPLE AUDIO/VIDEO CABLE TESTER

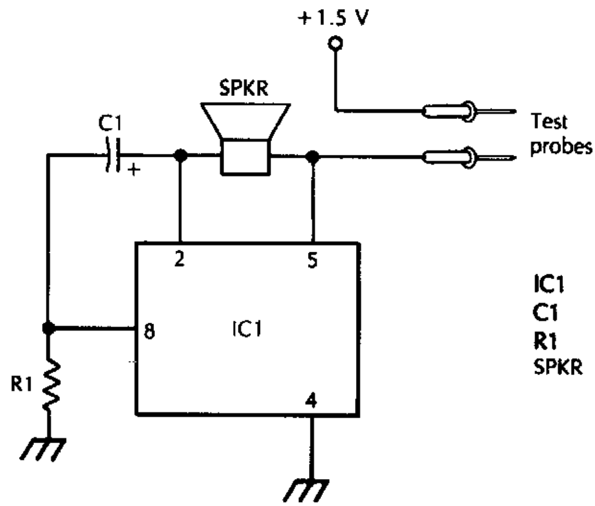


ELECTRONICS NOW ***Fig. 19-3***

As simple as it looks, the audio/video cable tester tests cables while they are plugged in. You can flex them vigorously while listening to the built-in buzzer.

Simply plug the ends of the cable in the appropriate jack. Only one end of the cable need be plugged in for a complete short test; the other end is left free. If the buzzer sounds, there is a short circuit somewhere in the cable. If nothing is heard, test for an intermittent short by flexing the cable several times, particularly in the plug area of both the free and plugged-in ends.

AUDIBLE CONTINUITY TESTER



IC1	**LM3909 LED flasher/oscillator IC**
C1	**10 µF 10 V electrolytic capacitor**
R1	**1 kΩ ¼ W 5% resistor**
SPKR	**Small 8Ω speaker**

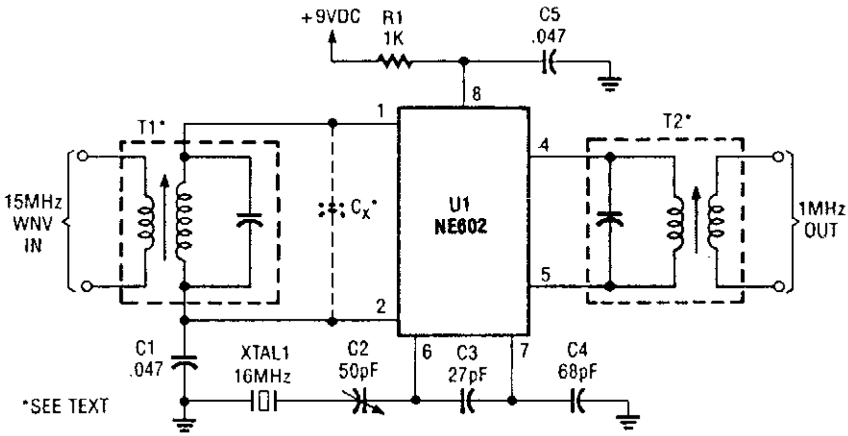
McGRAW-HILL ***Fig. 19-4***

20

Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

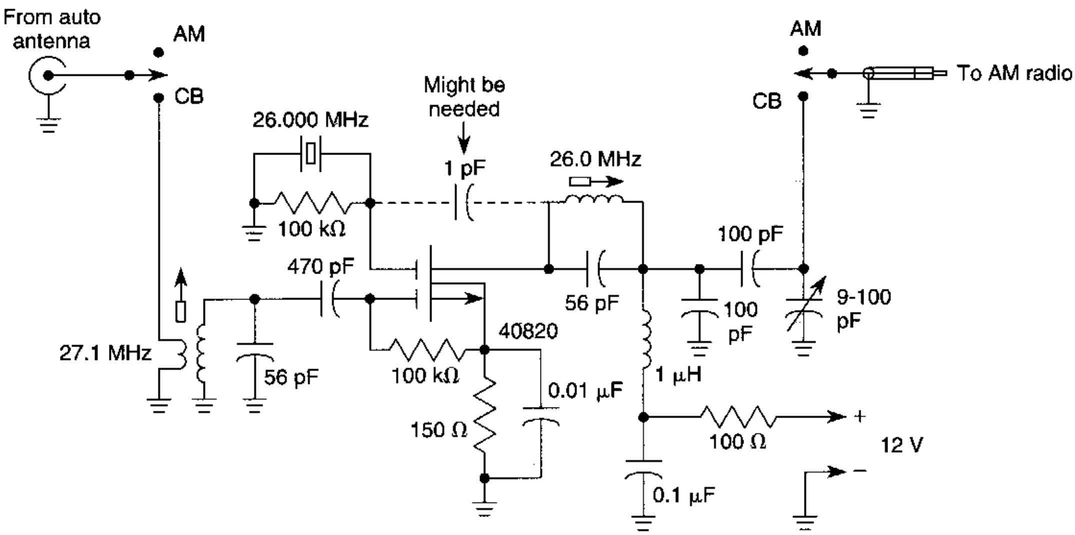
WWV CONVERTER



RADIO CRAFT *Fig. 20-1*

This converter heterodynes the 15-MHz WWV signal with a 16-MHz oscillator so that it can be heard at 1 MHz on an AM broadcast receiver. T1 and T2 are a modified 10.7-MHz IF transformer and AM BC oscillator coil, respectively.

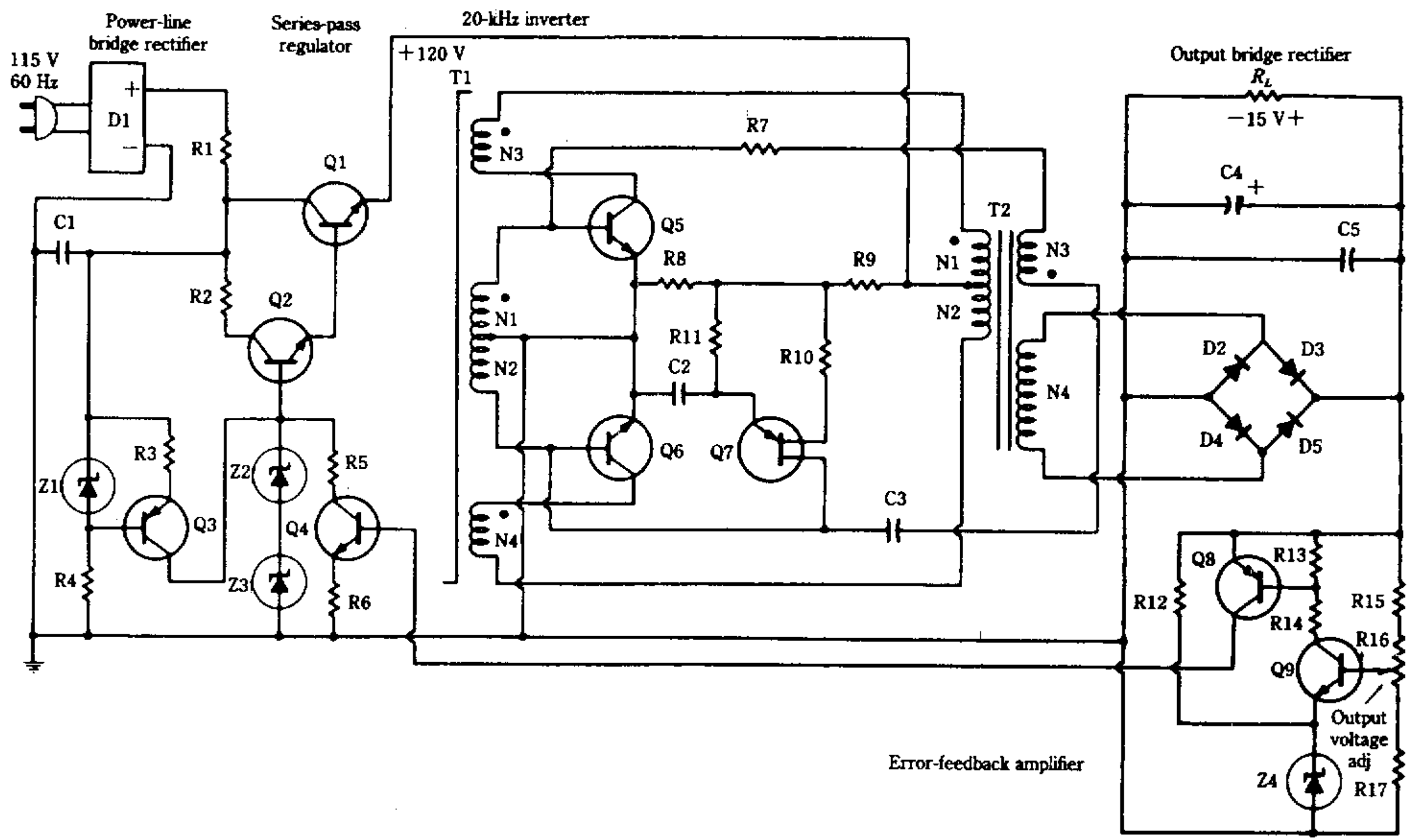
SIMPLE HF RECEIVE CONVERTER



WILLIAM SHEETS *Fig. 20-2*

Designed for CB reception, this crystal-controlled converter uses one 40820 dual-gate MOSFET. The circuit will work with any crystal either 3rd overtone or fundamental, over 1 to 50 MHz.

225-W, 15-V OUTPUT CONVERTER



McGRAW-HILL

Fig. 20-3

A converter designed to supply ±15 Vdc is shown. This converter is several times lighter in weight than an equivalent.

C1 — 2500 μF, 350 V Electrolyic
C2 — 0.1 μF Disc Ceramic
C3 — 0.1 μF Paper
C4 — 10 μF Electrolyic
C5 — 0.25 μF Paper
D1 — MDA-980-4 Bridge Rectifier Assembly
D2, D3, D4, D5 — 1N5826, 20 V 15 A

Q1, Q5, Q6, — 2N6307
Q2, Q4 — 2N5052
Q3 — 2N5345
Q7 — 2N4870
Q8 — 2N3905
Q9 — 2N3903

All Resistors in Ohms and 1/2 W Unless Otherwise Noted

R1 — 1, 10 W
R2 — 100
R3 — 82
R4 — 22K
R5 — 1.5K, 15 W
R6 — 200
R7 — 15
R8 — 4.7K
R9 — 51

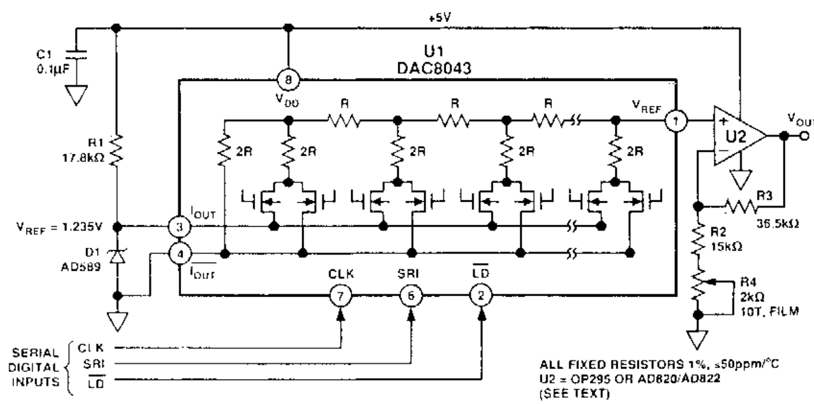
R10 — 1K
R11 — 10K
R12 — 270
R13 — 1K
R14 — 7.5K
R15 — 2.5K
R16 — 5K
R17 — 3.5K

T1 — Core — Magnetics Inc. 80623 — 1/2 D — 080
N1, N2 — 20 Turns Each, No. 30 AWG (Bifilar)
N3, N4 — 3 Turns Each, No. 20 AWG

T2 — Core — Arnold 6T 5800 D1
N1, N2 — 100 Turns Each, No. 20 AWG (Bifilar)
N3 — 7 Turns No. 26 AWG
N4 — 12 Turns Each, No. 12 AWG (No. 16 AWG, 3 in Parallel)

Z1 — 1N4733, 5.1 V
Z2, Z3 — 1N4760, 68 V
Z4 — 1N4736

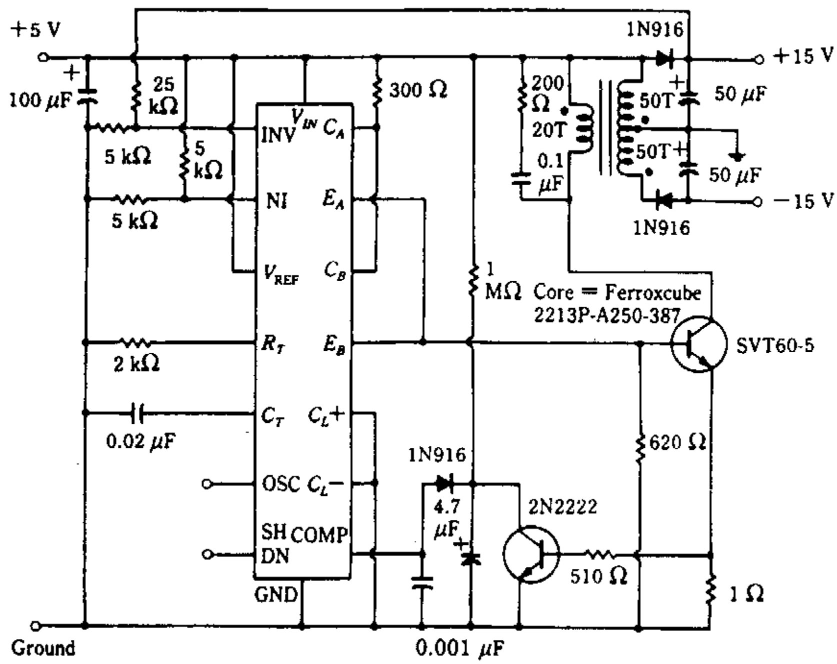
12-BIT DAC



ANALOG DEVICES ***Fig. 20-4***

This circuit uses an Analog Devices DAC-8043 12-bit multiplying DAC. The output voltage will be $D/4096 \times V_{ref}$ where D is the numerical value of the digital input word (0 to 4095). V_{ref} is 1.235 volts in this circuit.

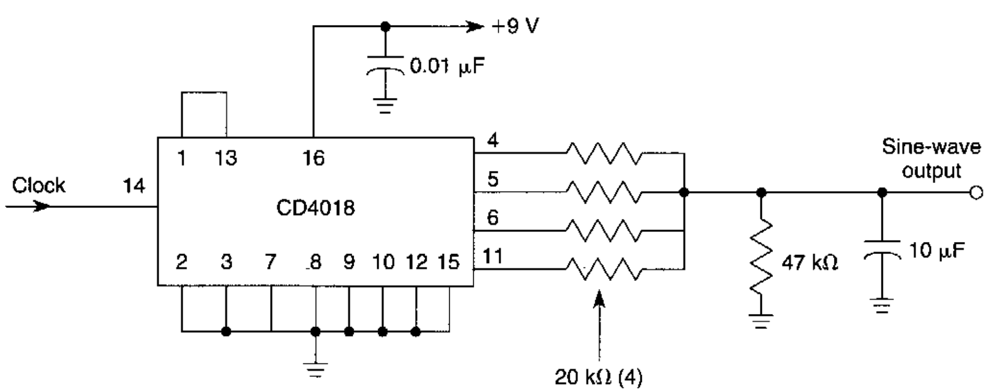
DRIVEN FLYBACK CONVERTER



McGRAW-HILL *Fig. 20-5*

This circuit uses an SG1524 Silicon General regulating pulse width modulator and provides ±15 V from a 5-V supply rail.

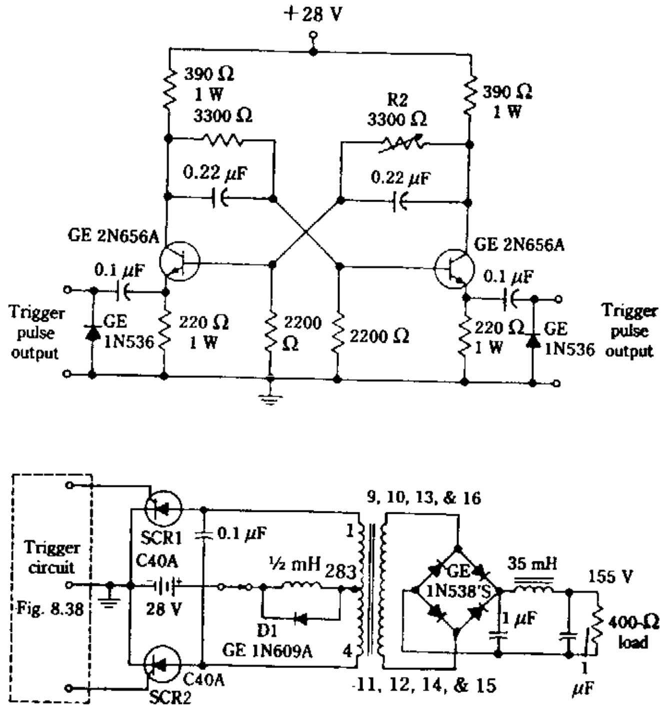
SINE-WAVE CONVERTER



WILLIAM SHEETS *Fig. 20-6*

This circuit produces a sine wave with a low-frequency clock input. The clock rate should be 100 Hz or less.

SCR CONVERTER

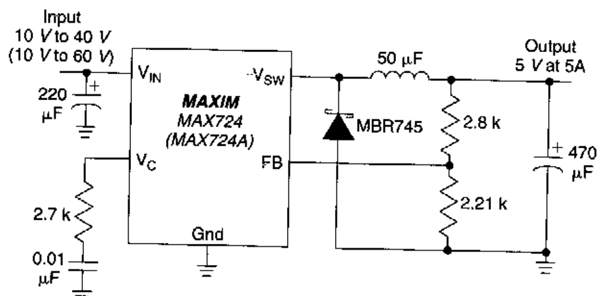


McGRAW-HILL

Fig. 20-7

Two SCR devices are used in a push-pull driver to convert 28 Vdc to 155 Vdc, using the transformer and bridge shown. A center-tapped transformer with 24 V:120 V could be used for 60-Hz applications. The trigger circuit supplies a push-pull drive signal.

5-V, 5-A STEP-DOWN CONVERTER



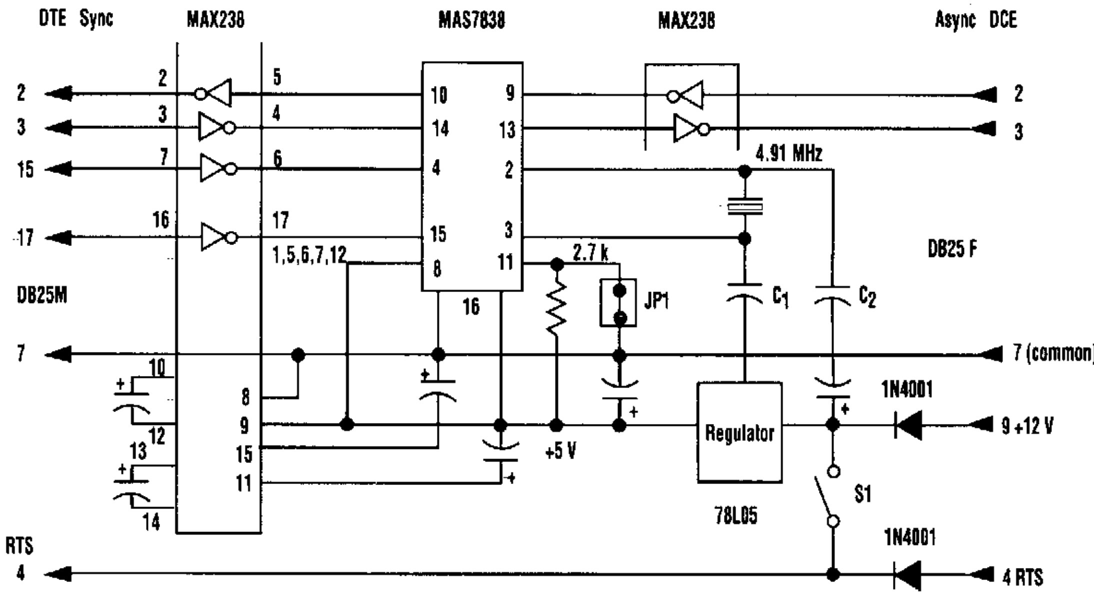
5A step-down converter

MAXIM

Fig. 20-8

This circuit is useful where power must be distributed by a higher (10 to 60 V) bus. The circuit reduces power dissipation and eliminates inefficient passive linear regulators. The switching frequency is in the 100-kHz region.

SYNC-TO-ASYNC CONVERTER



Point-topoint adapter wiring

1........1
5........5
6........6
7........7
8........8
9........9
10........10
15........15
17........17
20........20
21........21
24........24

Notes:

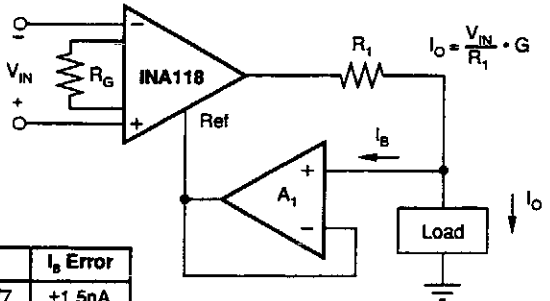
- Maxim RS232 interface: MAX238
- Sipex A/S converter: MAS7838
- Generic regulator: 78L05
- C_1 and C_2 = 20-pF disk-type; other capacitors (6) = 1 µF, 35 V
- Resistor = 1/8 W, 5%

ELECTRONIC DESIGN *Fig. 20-9*

This simple converter consists of two ICs and a voltage regulator. The Sipex MAS7838, which acts as the converter, selects the conversion speed to that of the synchronized data clock. It has internal switches and registers to perform the async-to-sync, or sync-to-async, conversion. The Maxim MAX238 provides the RS-232 drivers and receivers for interfacing with the data bus. These chips require a 5-Vdc power supply; a generic 78L05 reduces the +12 V at the DB25 pin 9 to the +5 V needed. A crystal frequency of 4.91 MHz is suitable for converting to 19.2 kbits or a sub-multiple (9.6, 4.8, 2.4, etc.). Two 1N4001 diodes protect the external RTS (ready to send) control circuitry if the RTS is enabled by S1. When JP1 is removed, the converter is transparent in the sync mode and no conversion will occur.

The completed unit is mounted atop a universal breakout adapter, and the control lines are jumpered according to the chart in the figure. The physical size is approximately 1 × 2.25 × 2.5 inches and will easily plug into the DB25 socket on a synchronized data communications equipment (DCE) communication device.

DIFFERENTIAL VOLTAGE-TO-CURRENT CONVERTER



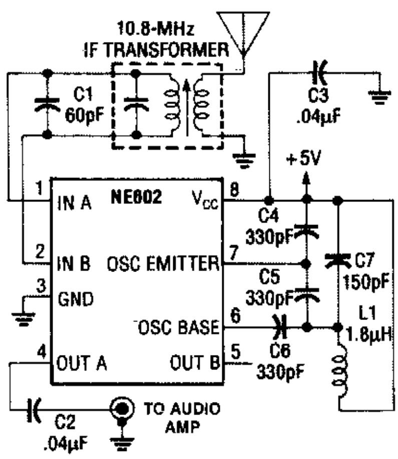
A_1	I_B Error
OPA177	±1.5nA
OPA602	±1pA
OPA128	±75fA

DESIRED GAIN	R_G (Ω)	NEAREST 1% R_G (Ω)
1	NC	NC
2	50.00k	49.9k
5	12.50k	12.4k
10	5.556k	5.62k
20	2.632k	2.61k
50	1.02k	1.02k
100	505.1	511
200	251.3	249
500	100.2	100
1000	50.05	49.9
2000	25.01	24.9
5000	10.00	10
10000	5.001	4.99

BURR-BROWN

Fig. 20-10

DIRECT-CONVERSION 7-MHz RECEIVER

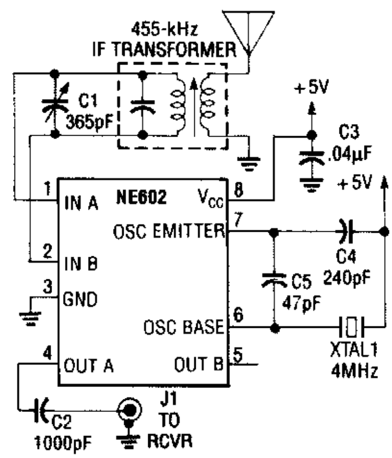


RADIO-ELECTRONICS

Fig. 20-11

An NE602 is used to mix signals in the 7-MHz range with an LO and to produce audio output.

LOW-FREQUENCY CONVERTER

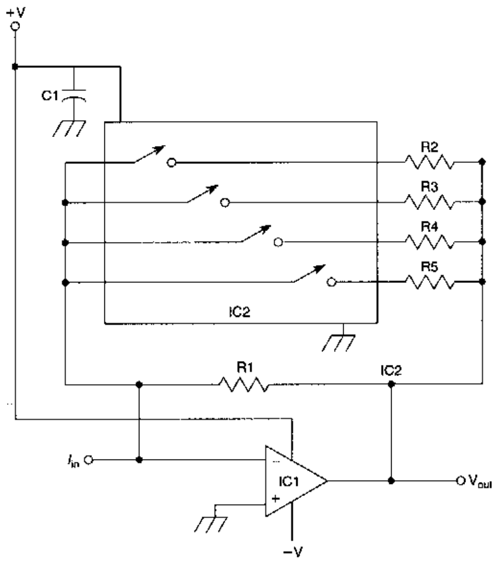


RADIO-ELECTRONICS

Fig. 20-12

This converter circuit translates the 350- to 500-kHz range to 4.35 to 4.50 MHz, enabling the frequency range to be received on a conventional shortwave receiver.

PROGRAMMABLE CURRENT-TO-VOLTAGE CONVERTER



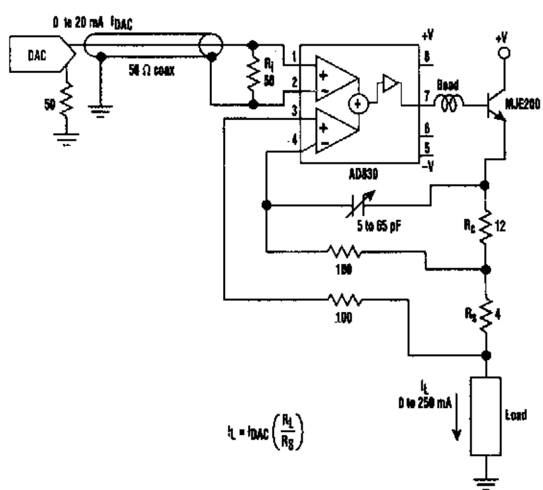
Programmable current-to-voltage converter permits you to electrically select from 16 resistor values using bilateral switches.

IC1	741 op amp (or similar)
IC2	CD4066 quad bilateral switch
C1	0.1-µF capacitor
R1	10-kΩ, ¼-W 5% resistor
R2	4.7-kΩ, ¼-W 5% resistor
R3	2.2-kΩ, ¼-W 5% resistor
R4	1.2-kΩ, ¼-W 5% resistor
R5	100-Ω, ¼-W 5% resistor

McGRAW-HILL

Fig. 20-13

CURRENT-TO-VOLTAGE CONVERTER WITH BOOST TRANSISTOR

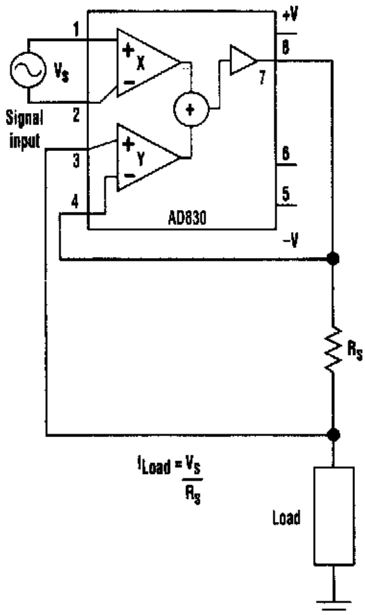


ELECTRONIC DESIGN

Fig. 20-14

A transistor such as the MJE200 can be added to an Analog Devices AD830 to produce this current to voltage converter. Loads to 250 mA can be driven. The 5- to 65-pF trimmer is for compensation.

CURRENT-TO-VOLTAGE CONVERTER FOR GROUNDED LOADS



This circuit uses an Analog Devices AD830 video difference amplifier. The circuit consists of two differential inputs. Unlike a conventional op amp, the AD830's output is nulled when the sum of the differences of the two inputs is zero.

The AD830's stated unity-gain bandwidth is 60 MHz, and the device is capable of driving up to ±30 mA directly. The differential input voltage is limited to ±2 V, while the maximum power supply is ±15 V.

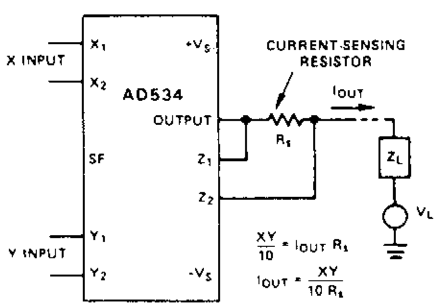
If more output current is desired, the AD830 can drive a bipolar transistor (such as an MJE200) directly. This will produce a one-sided output.

A ferrite bead can be placed on the base to prevent oscillation under some conditions. Compensation can be added by splitting R_S and adding a variable capacitor. A resistor can be positioned at the input to match the amplifier's input to a transmission line.

ELECTRONIC DESIGN

Fig. 20-15

OUTPUT-TO-CURRENT CONVERTER



Occasionally, it is preferable to generate a current, rather than a voltage, output into the load. The availability of differential inputs allows this to be accomplished in any of the four basic modes.

If the output is to integrated, Z_L can be simple high-quality capacitor, unloaded by an op amp connected as a high-impedance follower. Note that, if desired, one side of a rest switch can be grounded.

The compliance constraint for this configuration, where V_L is an arbitrary common-mode potential, is:

$$|V_L + I_{OUT}(Z_L + R_S)| \le 12 \text{ V}$$

ANALOG DEVICES

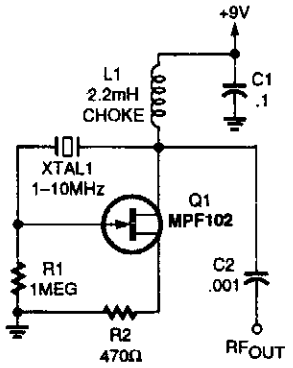
Fig. 20-16

21

Crystal Oscillator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

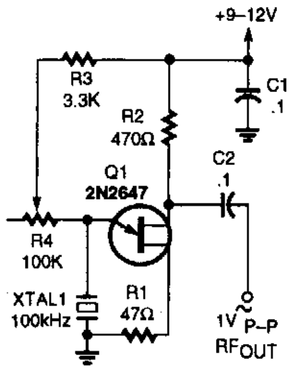
FET QUARTZ CRYSTAL OSCILLATOR



POPULAR ELECTRONICS *Fig. 21-1*

This oscillator uses an MPF102 JFET as an active element.

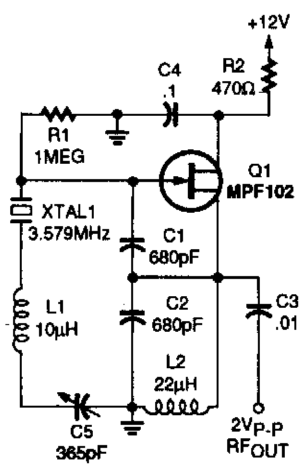
CRYSTAL OSCILLATOR I



POPULAR ELECTRONICS *Fig. 21-2*

In this circuit, series-resonant crystal XTAL1 is used as a frequency-determining element. XTAL1 is between 0.1 to 10 MHz.

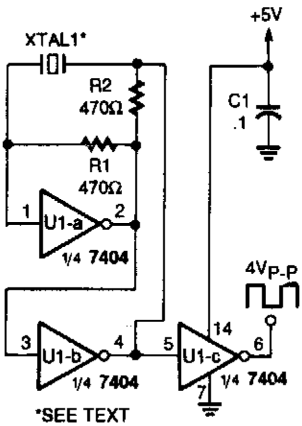
FET VXO CIRCUIT



POPULAR ELECTRONICS *Fig. 21-3*

An MPF 102 is used in a Colpitts-type oscillator in order to pull the crystal frequency slightly.

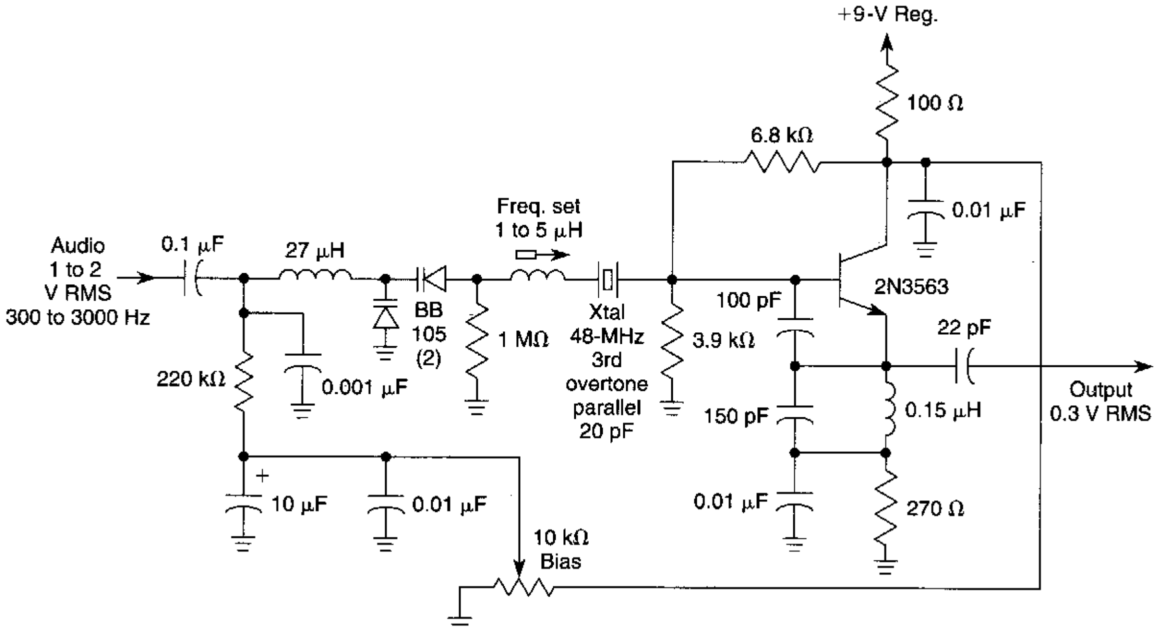
UJT 100-kHz CALIBRATION OSCILLATOR



POPULAR ELECTRONICS *Fig. 21-4*

This unusual 100-kHz oscillator (whose frequency is determined by XTAL1) can be used as a marker generator to calibrate the analog dial of a communication receiver, or its output can be fed to a divider counter to produce a stable lower-frequency output for use as a clock-signal generator.

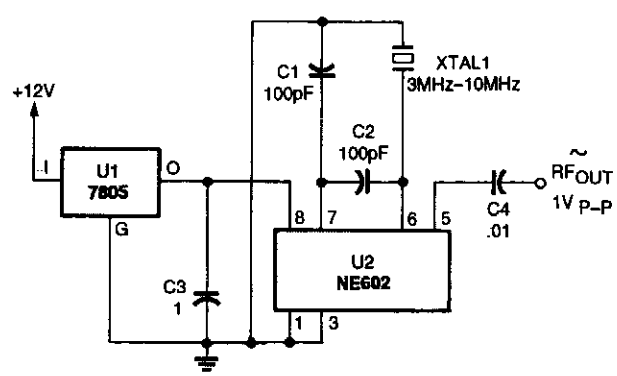
CRYSTAL OSCILLATOR WITH FM CAPABILITY



WILLIAM SHEETS **Fig. 21-5**

This crystal oscillator produces a good FM signal that can be tripled to 146 MHz and produces a clean 5-kHz deviation signal for FM voice. The bias control is adjusted for cleanest audio while the 1- to 5-μH coil is adjusted to set the oscillator frequency to the exact setting required.

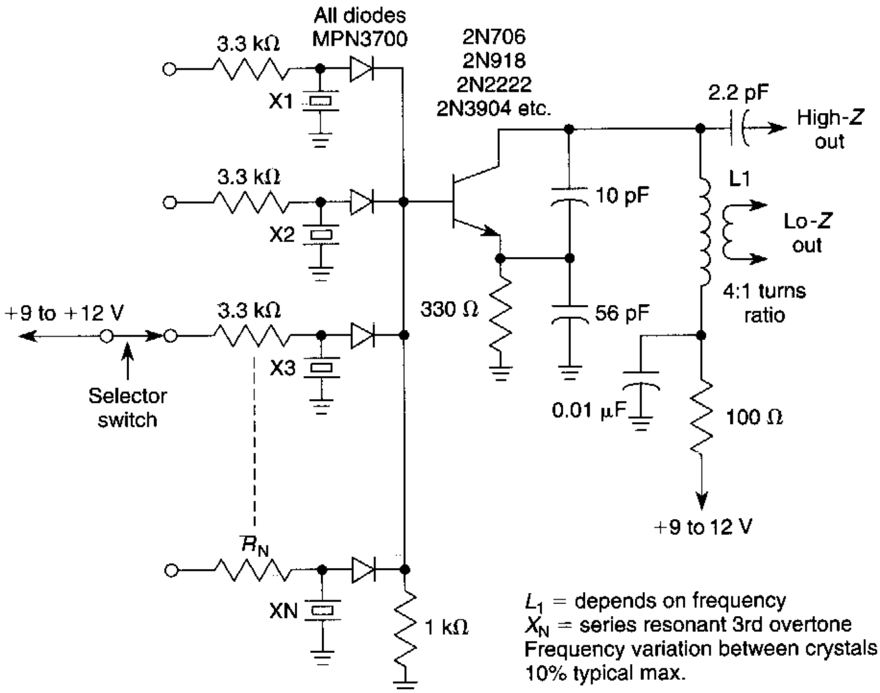
CRYSTAL OSCILLATOR II



POPULAR ELECTRONICS **Fig. 21-6**

An NE602 can be used as a crystal oscillator.

dc-SWITCHED CRYSTAL OSCILLATOR

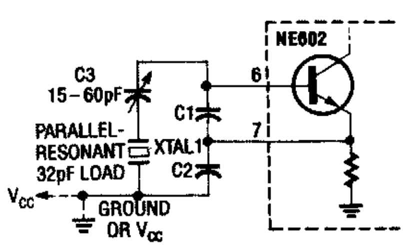


WILLIAM SHEETS

Fig. 21-7

This circuit is useful where several different crystal frequencies must be switched using a dc source. The values shown are typical for 40- to 60-MHz third-overtone crystals. Limitation on number of crystals depends on PIN diode capacitance and layout factors, but up to 5 or 10 crystals is possible.

CRYSTAL OSCILLATOR WITH ADJUSTABLE FREQUENCY

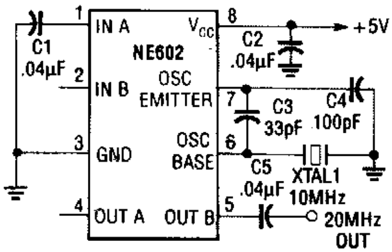


RADIO-ELECTRONICS

Fig. 21-8

In this crystal oscillator circuit, C3 adjusts the frequency of the oscillator for exact netting. The crystal is a fundamental type. C_1 = 100 pF and C_2 = 1000 pF are typical.

FREQUENCY DOUBLER AND CRYSTAL OSCILLATOR

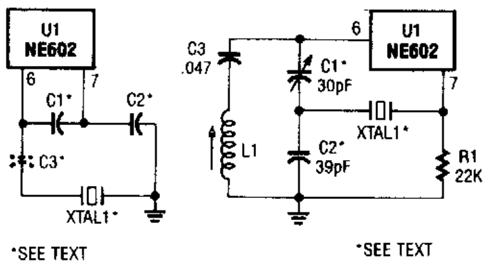


This frequency doubler produces a sine wave at twice the frequency of XTAL1. Notice that the output is taken only from OUT B (pin 5), while OUT A (pin 4) is left open.

RADIO-ELECTRONICS

Fig. 21-9

CRYSTAL OSCILLATOR III



These circuits are for use with a crystal-controlled LO using the NE602. C1, C2, and C3 are for crystals in the 5-MHz region and are approximately chosen from

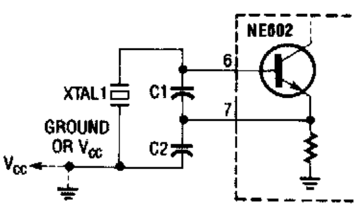
$$C_1 = \frac{100}{\sqrt{f_{MHz}}}\ \text{pF},\ C_2 = \frac{1000}{\sqrt{f_{MHz}}}\ \text{pF}$$

C3 is for fine tuning the crystal frequency and will be 20 to 50 pF typically.

RADIO CRAFT

Fig. 21-10

COLPITTS OSCILLATOR



f MHz	C1=100pF/√f	C2=1000pF/f
1	100	1000
2	70	500
4	50	250
10	32	100
20	22	50

Here: $L_1 = 7\ \mu H/f$, $C_1 \approx C_2 \approx C_3 \approx 2400\ pF/f$, where f is in MHz. In this circuit, the oscillator is free-running.

RADIO-ELECTRONICS

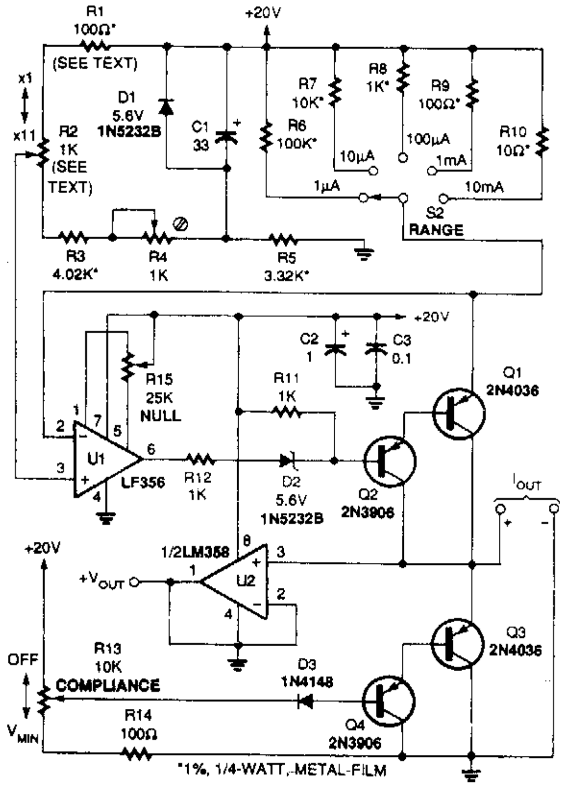
Fig. 21-11

22

Current Source and Sink Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

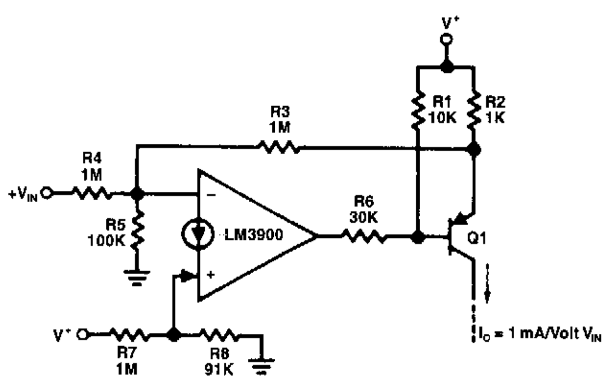
CURRENT GENERATOR



POPULAR ELECTRONICS *Fig. 22-1*

This circuit is useful for supplying constant current to test semiconductors. V_{OUT} from U2 reads the voltage across the load connected to I_{OUT}. R13 adjusts the supply compliance from 1 to about 18 V.

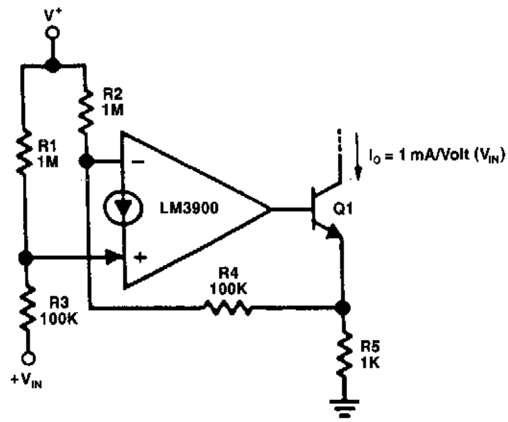
VOLTAGE-CONTROLLED CURRENT SOURCE



NATIONAL SEMICONDUCTOR *Fig. 22-2*

A voltage-variable current source is shown in the figure. The transconductance is $-(1/R_2)$ as the voltage gain from the input terminal to the emitter of Q1 is −1. For V_{in} = 0 Vdc, the output current is essentially 0 mA dc. Resistors R1 and R6 guarantee that the amplifier can turn OFF transistor Q1.

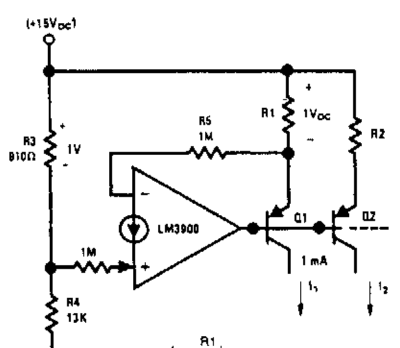
VOLTAGE-CONTROLLED CURRENT SINK



A voltage-variable current sink is shown in the figure. The output current is 1 mA per volt of V_{in} (as R_5 = 1 kΩ and the gain is +1). This circuit provides approximately 0 mA output current for V_{in} = 0 V DCL.

NATIONAL SEMICONDUCTOR *Fig. 22-3*

MULTIPLE FIXED CURRENT SOURCE



A multiple fixed current source is provided by the circuit. A reference voltage (1 Vdc) is established across resistor R3 by the resistive divider (R3 and R4). Negative feedback is used to cause the voltage drop across R1 to also be 1 Vdc. This controls the emitter current of transistor Q1 and if we neglect the small current diverted into the (–) input via the 1-MΩ input resistor (13.5 μA) and the base current of Q1 and Q2 (an additional 2% loss if the β of these transistors is 100), essentially this same current is available out of the collector of Q1.

Larger input resistors can be used to reduce current loss and a Darlington connection can be used to reduce errors caused by the β of Q1.

The resistor, R2, can be used to scale the collector current of Q2 either above or below the 1-mA reference value.

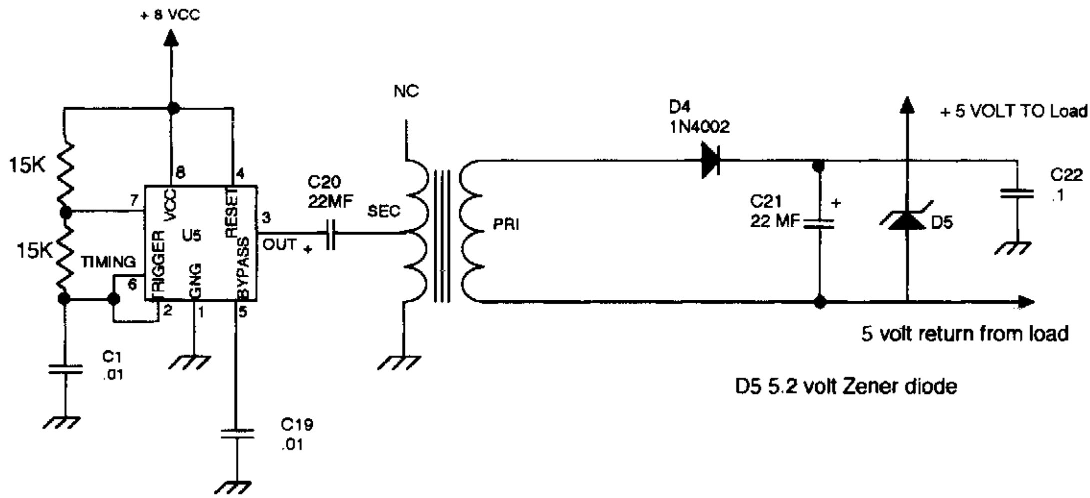
NATIONAL SEMICONDUCTOR *Fig. 22-4*

23

dc-to-dc Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

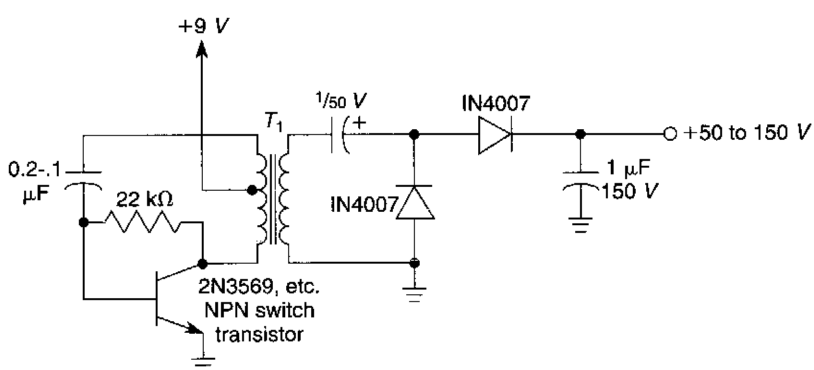
ISOLATED dc-to-dc CONVERTER



73 AMATEUR RADIO TODAY *Fig. 23-1*

A NE555 timer is used to drive a small transformer to change the 5- to 7-Vp-p output of the NE555 to a suitable value to drive a rectifier/Zener combination. This method is useful where a small isolated power source is needed.

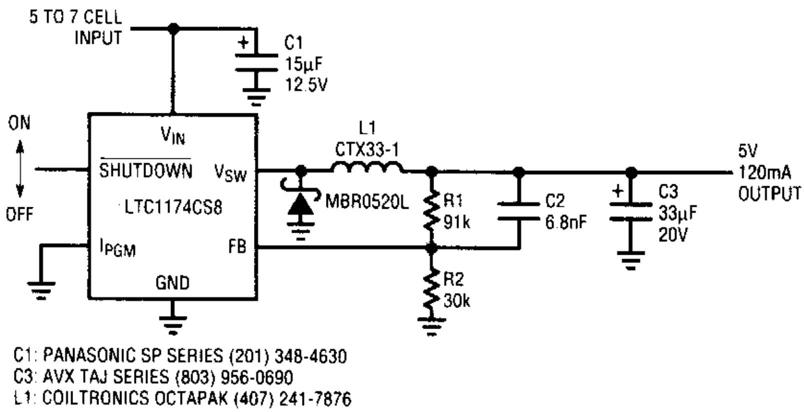
dc-to-dc CONVERTER



WILLIAM SHEETS *Fig. 23-2*

This converter should be useful where a few milliamperes of dc at a higher voltage than available supplies can deliver is needed. T1 is typically a 1 kΩ CT:10-kΩ transistor audio transformer. Depending on T1, about 50 to 150 Vdc can be obtained at a few milliamperes.

ULTRA LOW-POWER dc-to-dc CONVERTER FOR PERSONAL COMMUNICATIONS PRODUCTS

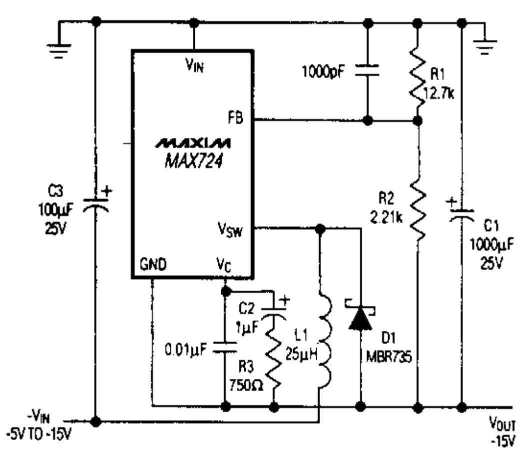


LINEAR TECHNOLOGY *Fig. 23-3*

The LTC1174 step-down converter is designed specifically to eliminate noise at audio frequencies while maintaining high efficiency at low output currents. This circuit shows a 5-V, 120-mA output derived from 5 to 7 NiCad or NiMH cells. Small input and output capacitors that are capable of handling the necessary ripple currents help conserve space. In applications where shutdown is desired, this feature is available (otherwise short this pin to V_{in}).

The LTC1174's internal switch, connected between V_{in} and V_{sw}, is current controlled at a peak of approximately 340 mA. Low peak switch current is one of the key features that allows the LTC1174 to minimize system noise compared to other chips that carry significantly higher peak currents, easing shielding and filtering requirements, and decreasing component stresses. Output current of up to 450 mA is possible with this device by connecting the 1 pgm pin to V_{in}. This increases the peak current to 600 mA, allowing for a high average output current.

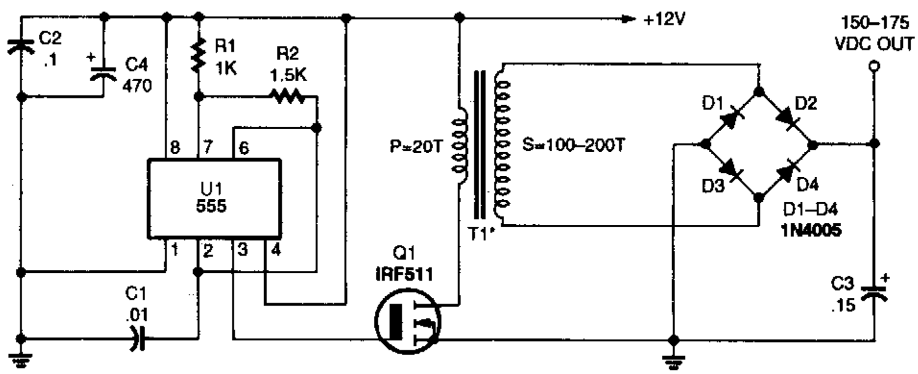
NEGATIVE STEP-UP dc-to-dc CONVERTER



A Maxim MAX724 is used in a step-up switching converter to provide −15-V output from a −5- to −15-V input.

MAXIM *Fig. 23-4*

dc-to-dc CONVERTER II



POPULAR ELECTRONICS *Fig. 23-5*

In this dc-to-dc converter, the 555 is used to produce a rising and collapsing field in T1's primary, generating a high voltage in T1's secondary winding. That voltage is then full-wave, bridge rectified by D1 through D4, and filtered by C3. T1 is an Amidon Associates EA-775-375E core and nylon bobbin, with #26 wire for the primary and #28 or #30 for the secondary. About 5 W of power is available.

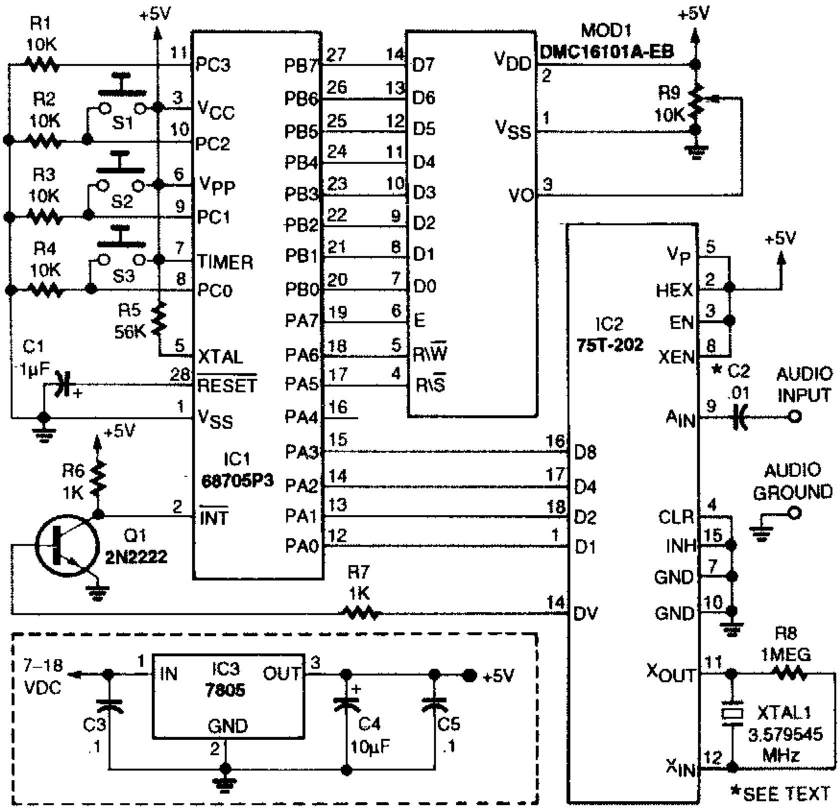
24

Decoder Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

DTMF Decoder
FM Stereo Decoder
Typical NE567 Tone Decoder Circuit
Video Line Decoder I
Video Line Decoder II

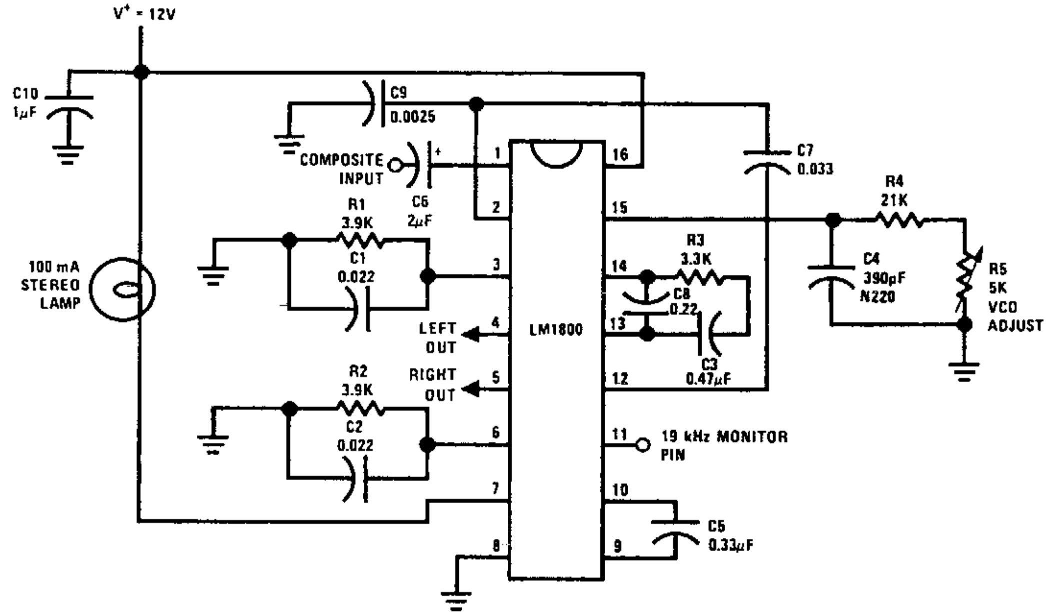
DTMF DECODER



ELECTRONICS NOW *Fig. 24-1*

This DTMF decoder uses a Motorola 68705P3 microcontroller and a 75T202 DTMF receiver (Silicon Systems, Inc.). An LCD module is used for the display (MOD1). Switch S1 is used to scroll the display, S2 clears the display, and S3 clears the memory.

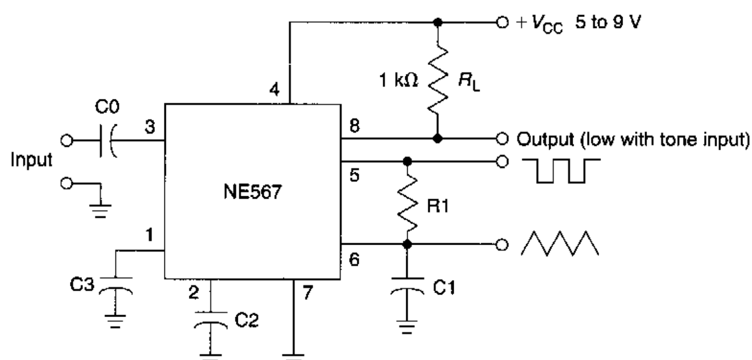
FM STEREO DECODER



NATIONAL SEMICONDUCTOR ***Fig. 24-2***

Using an LM1800, this circuit takes composite baseband MPX input and recovers L7R audio channels. The VCO is set for 19 kHz (or 15.7 kHz for TV applications) or as needed.

TYPICAL NE567 TONE DECODER CIRCUIT



$f_o = 1/1.1R_1C_1$ $BW = 1070\sqrt{V_i/f_oC_2}$ $V_1 \leq 200$ mV RMS

$X_{co} \approx < 2\ \text{k}\Omega$ @ f_o f_o Hz $C_3 \approx 2C_2$

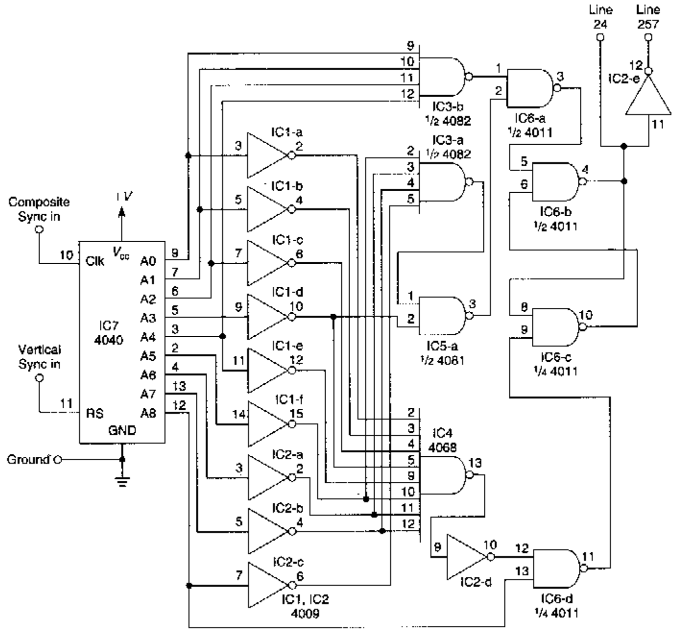
C_2 μF

V_1 V

WILLIAM SHEETS ***Fig. 24-3***

This circuit illustrates use of NE567 as a tone decoder.

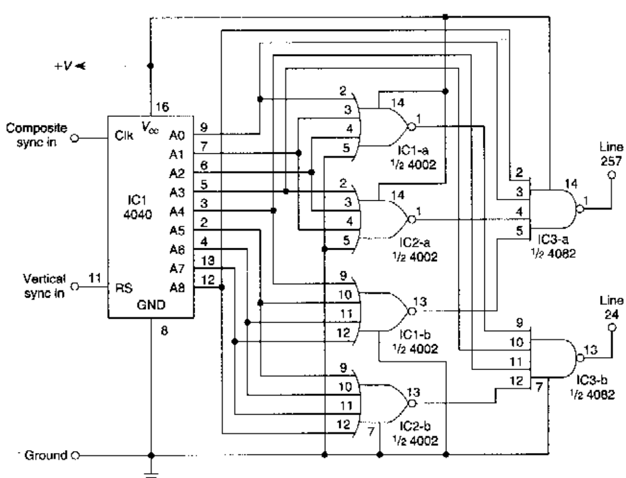
VIDEO LINE DECODER I



ELECTRONICS NOW ***Fig. 24-4***

This line decoder uses only one pin for the line indicator.

VIDEO LINE DECODER II



ELECTRONICS NOW ***Fig. 24-5***

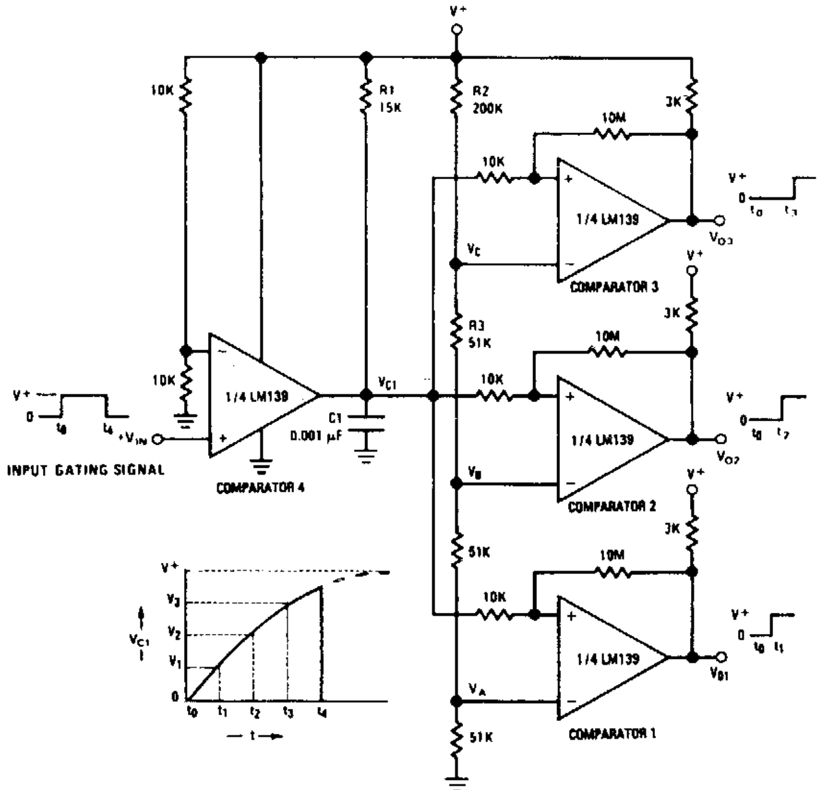
This circuit will produce pulses useful for gating lines 24 and 257 of a video signal, but by changing the decoding logic, other lines can be decoded.

25

Delay Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

TIME-DELAY GENERATOR

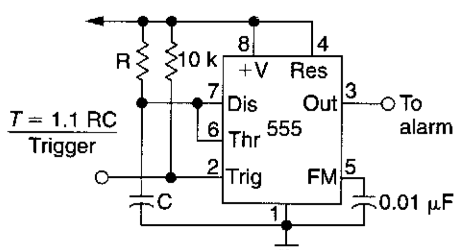


NATIONAL SEMICONDUCTOR

Fig. 25-1

This circuit uses a charging capacitor and three comparators to read the voltage across it.

SIMPLE TIME DELAY CIRCUIT



ELECTRONICS NOW

Fig. 25-2

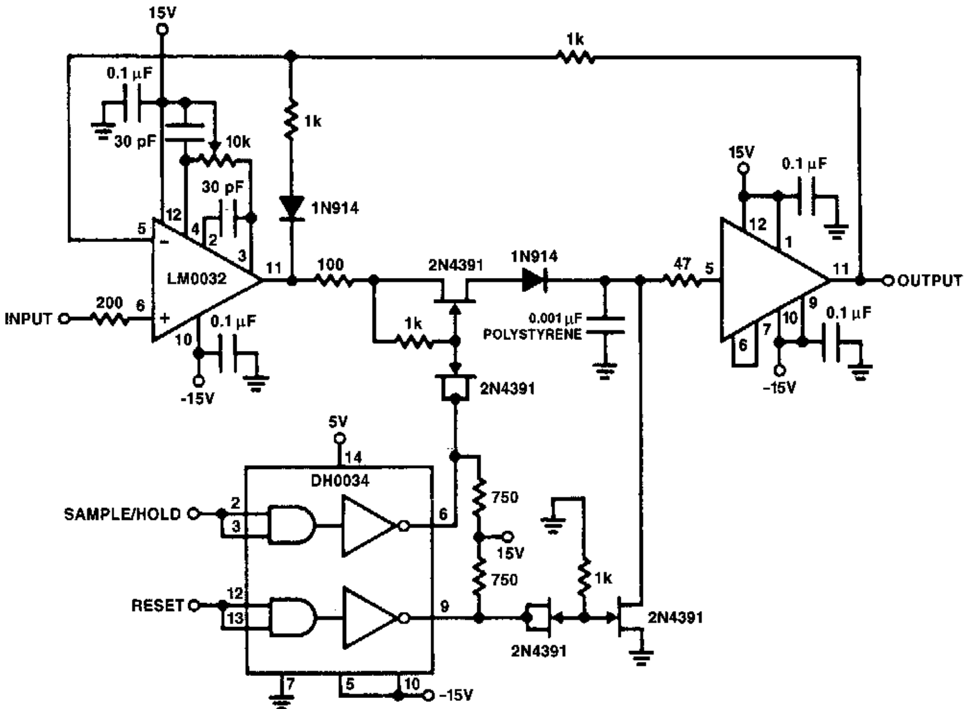
Rotating the potentiometer wiper will change the time delay from the 555 IC.

26

Detector Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

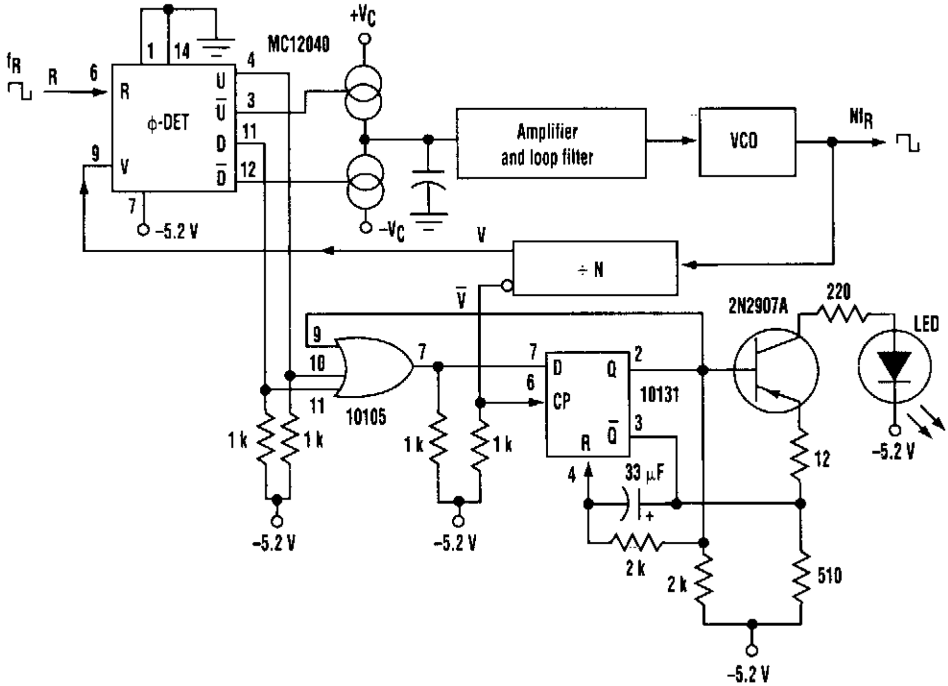
HIGH-SPEED PEAK DETECTOR WITH HOLD AND RESET CONTROLS

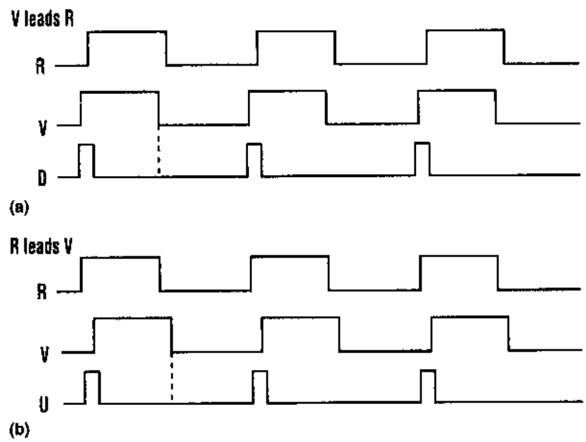


NATIONAL SEMICONDUCTOR *Fig. 26-1*

The LH0033 and LH0063 are useful in high-speed sample-and-hold or peak detector circuits because of their very high speed and low-bias-current FET input stages. The high-speed peak detector circuit shown could be changed to a sample-and-hold circuit simply by removing the detector diode and reset circuitry. For best accuracy, the circuit can be trimmed with the 10-kΩ offset adjustment pot shown. The circuit has a typical acquisition time of 900 ns, to 0.1% of the final value for the 10-V input step signal, and a droop rate of 100 μV/ms. Even faster acquisition time can be achieved by reducing the hold capacitor value.

LOCK DETECTOR





ELECTRONIC DESIGN

Fig. 26-2

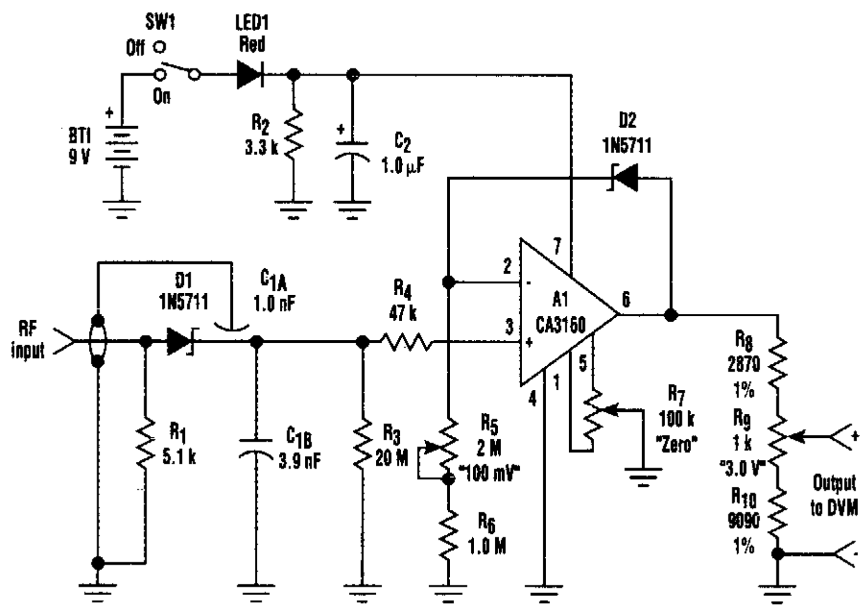
LOCK DETECTOR (*Cont.*)

This PLL lock indicator not only can detect a "locked" or "off-of-lock" condition, but also even if a single pulse or transition has been missed.

When being sampled by the flip-flop, if the *V* signal leads the input reference signal *R* and the rising edge of *R* is lost, the *D* signal will remain high throughout the interval, allowing the flip-flop to be clocked high (Fig. 26-2A). If the *R* signal leads the *V* signal when the transition is missed, the rising edge of the *V* signal will trigger the *D* signal of the phase detector, causing the LED to blink (Fig. 26-2B).

A "lock" detector often is used with a phase-locked loop (PLL) or synthesizer to indicate when the loop is phase-locked with an input signal. This circuit can be helpful, but single cycle skips usually will go undetected because of the presence of the low-pass filter.

LINEARIZED RF DETECTOR

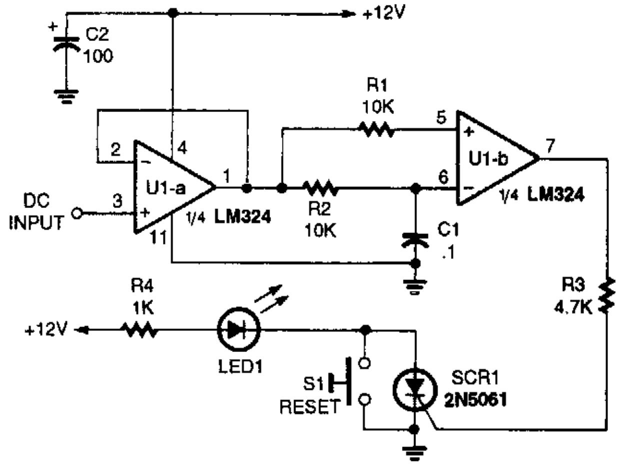


Input range: 100 mV rms to 80 V rms (2%)

ELECTRONIC DESIGN *Fig. 26-3*

The circuit produces an extremely linear dc output for RF inputs between 80 mV rms and 4.0 mV rms. For inputs below 50 V rms, the dc output quickly drops to 0 V.

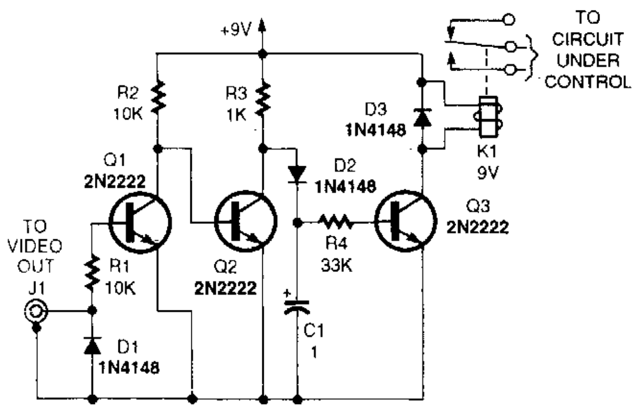
GLITCH DETECTOR



POPULAR ELECTRONICS *Fig. 26-4*

In the circuit, two op amps (half of an LM324 quad op amp) and an SCR are direct coupled in a dc-voltage monitoring circuit. Op-amp U1-a is configured as a voltage follower, which feeds the bridged inputs of the second op amp, U1-b. A resistor/capacitor combination (R2/C1) connected to the negative input of UJ1-b forms an RC time-delay circuit. As long as there is no change in the dc-voltage level at either of U1-b's inputs, its output is near zero. If a voltage glitch occurs, the RC timing circuit will delay the voltage change at the op amp's inverting input, causing its output to go high, triggering SCR1 and causing LED1 to light. The circuit's sensitivity allows it to detect voltage changes in the millivolt range. Pressing S1 diverts the SCR's holding current to ground, causing it to turn off and reset the circuit.

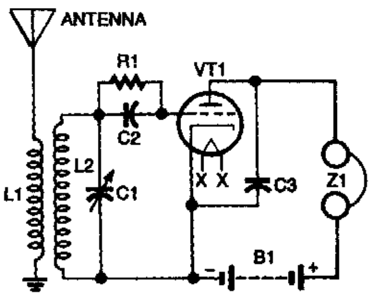
VCR VIDEO DETECTOR CONTROLLER



POPULAR ELECTRONICS *Fig. 26-5*

This circuit uses the video output from a VCR or camera to control a relay. Video turns on Q1, cutting off Q2, allowing Q3 to be forward biased, activating relay K1. You can use the timer in your VCR and this unit to generate long time delays as well.

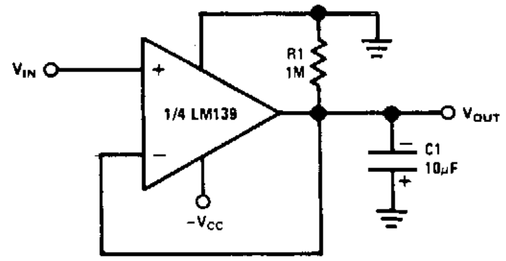
GRID-LEAK DETECTOR



ELECTRONICS NOW *Fig. 26-6*

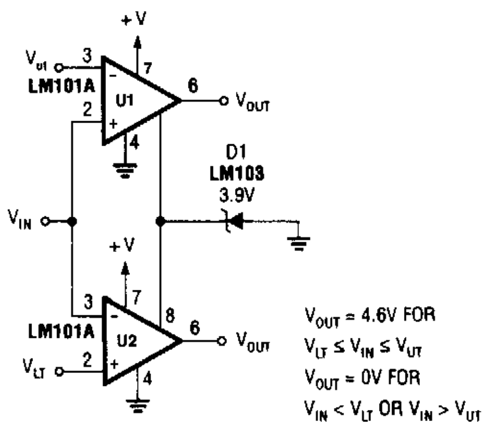
Tuned-circuit receiver with grid-leak detection.

NEGATIVE PEAK DETECTOR



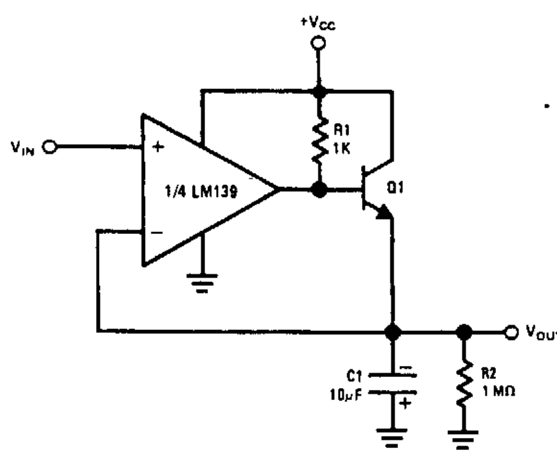
NATIONAL SEMICONDUCTOR *Fig. 26-7*

DOUBLE-ENDED LIMIT DETECTOR



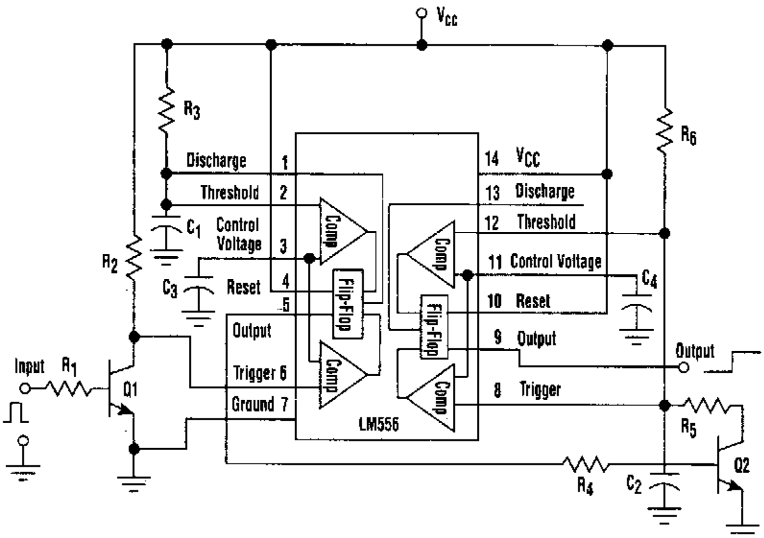
POPULAR ELECTRONICS *Fig. 26-8*

POSITIVE PEAK DETECTOR



NATIONAL SEMICONDUCTOR *Fig. 26-9*

LM556 FREQUENCY DETECTOR



ELECTRONIC DESIGN *Fig. 26-10*

The circuit (see the figure), is built around an LM556 dual-timer IC. The 556's first timer is wired as a one-shot and is used to stretch the incoming pulses into fixed-length pulses. The second timer, which is similar to an astable multivibrator (pin 13 remains disconnected), functions as follows:

The one-shot's fixed-length pulses, which are output on pin 5, turn on Q1 and discharge C2 through R5. If the frequency of the pulses is high enough, the voltage on C2 will fall below ⅓ VCC and the second timer's output, pin 9, will go to a logic 1. Conversely, if the frequency is low enough or is zero, the voltage on C2 will charge through R6 to a level above ⅔ VCC, and the pin 9 output will go to a logic "0."

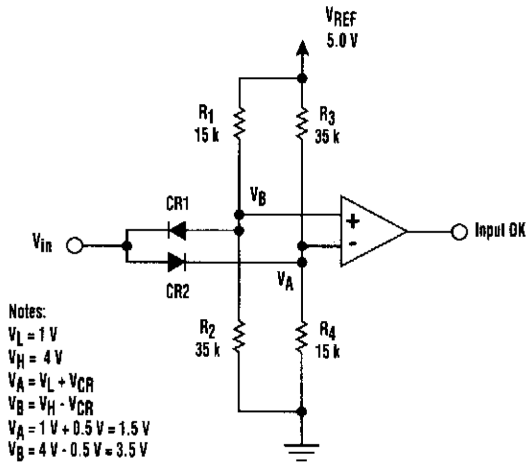
The idea is to keep the upper and lower peak voltage on C2 below ⅔ VCC and ⅓ VCC, respectively for a logic 1, and above ⅔ VCC and ⅓ VCC, respectively, for a logic 0.

To find the one-shot values, R3 and C1, select a pulse width ($1.1 \times R_3 \times C_1$) that's greater than the largest input pulse width and less than twice the inverse of the highest input frequency. To find R5, R6, and C2, first determine the duty cycle (t_{on}/t_{off}) of the input signal. Next, choose a standard value for C2 and calculate R6:

$$R_6 = \frac{[(t_{off} \times 0.61)^2 + t_{off}]}{C_2}$$

Also, $R_5 = R_6\ (t_{on}/t_{off})$. A tweak of resistors R5 and R6 might be needed to get the preferred response. Input signals with low duty cycles work the best. Finally, notice that capacitors C3 and C4 can be any value between 0.01 and 0.1 μF.

SINGLE COMPARATOR WINDOW DETECTOR



ELECTRONIC DESIGN

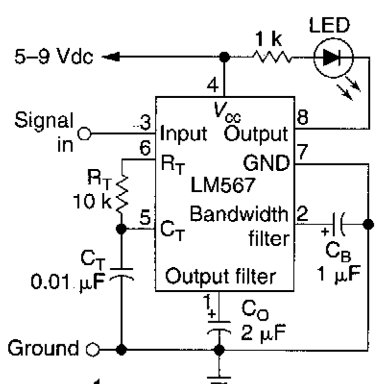
Fig. 26-11

Simply by adding two steering diodes, a window detector can be built using only a single comparator. The detector performs well for windows of about 1 V or greater, but it isn't suitable where extreme precision is required because the forward drops of the diodes vary.

In the basic circuit, two resistive dividers set threshold voltage levels at both the inverting and noninverting inputs of the comparator by dividing the reference voltage. The input voltage is steered to the appropriate comparator input by diodes CR1 and CR2.

When the input voltage is within the window, neither diode conducts, and the comparator is biased for a high output. When the input goes above the window, CR2 conducts and pulls the inverting input high, causing the comparator output to go low. When the input voltage goes below the window, CR1 conducts, pulling the noninverting input low, again causing the comparator output to go low. The source resistance of V_{in} must be low compared to the equivalent parallel resistance of each divider.

15-kHz TONE DETECTOR



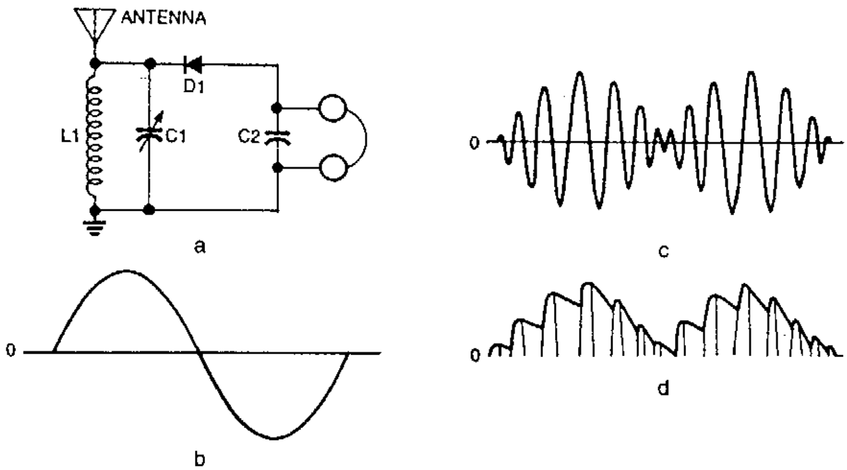
$$F = \frac{1}{1.1\,R_T\,C_T}$$

ELECTRONICS NOW!

Fig. 26-12

This circuit detects the presence of a 15-kHz audio signal and light the LED when it does so.

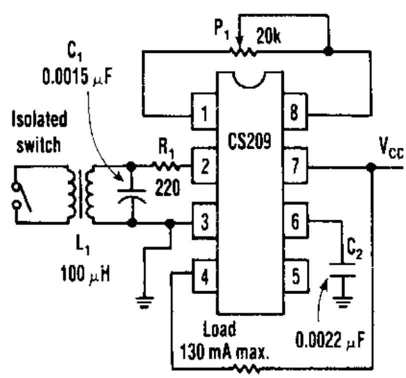
CRYSTAL RADIO DETECTOR



ELECTRONICS NOW! *Fig. 26-13*

This is a crystal detector receiver with headphones (Fig. 26-13A), audio-frequency signal (Fig. 26-13B), modulated signal (Fig. 26-13C), and a demodulated wave (Fig. 26-13D).

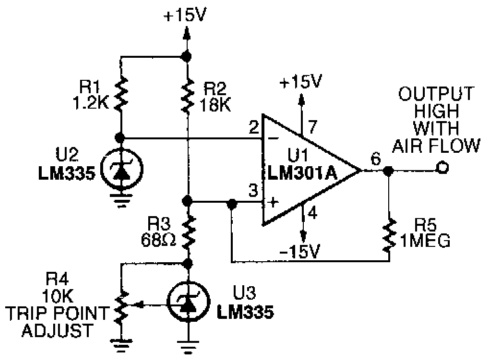
SWITCH CLOSURE CIRCUIT



ELECTRONIC DESIGN *Fig. 26-14*

A standard proximity detector circuit for the Cherry Semiconductor CS209 IC can detect an isolated switch closure by adding a few turns of wire around the circuit's inductor (Radio Shack 273-102). Moreover, the technique doesn't require any isolated power (see the figure). With the switch open, the potentiometer P1 is adjusted until the output switches off. When the switch is closed, the Q of the circuit changes and the output turns on. Capacitor C1 should be silvered mica, and potentiometer P1 should be a multiturn type such as the Bourns 3006P-1-203. A 9-V supply can be used for Vcc.

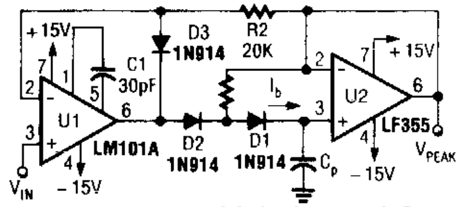
AIR FLOW DETECTOR



*SELF HEATING IS USED TO DETECT AIR FLOW

POPULAR ELECTRONICS *Fig. 26-15*

LOW DRIFT PEAK DETECTOR



BY ADDING D1 AND R_f,V_{D1}-0 DURING HOLD MODE. LEAKAGE OF D2 PROVIDED BY FEEDBACK PATH THROUGH R_f.

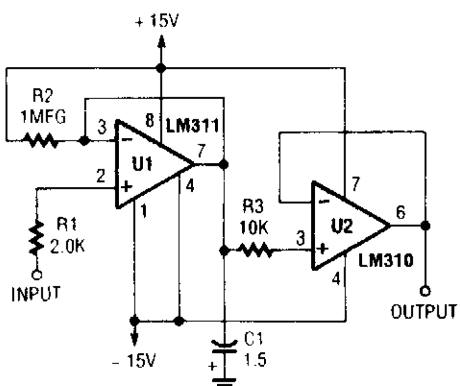
LEAKAGE OF CIRCUIT IS ESSENTIALLY I_b(LF155,LF156) PLUS CAPACITOR LEAKAGE OF Cp.

DIODE D3 CLAMPS V_{OUT} (A1) TO V_{IN}-V_{D3} TO IMPROVE SPEED AND TO LIMIT REVERSE BIAS OF D2.

MAXIMUM INPUT FREQUENCY SHOULD BE << $1/2\pi R_f C_{D2}$ WHERE C_{D2} IS SHUNT CAPACITANCE OF D2.

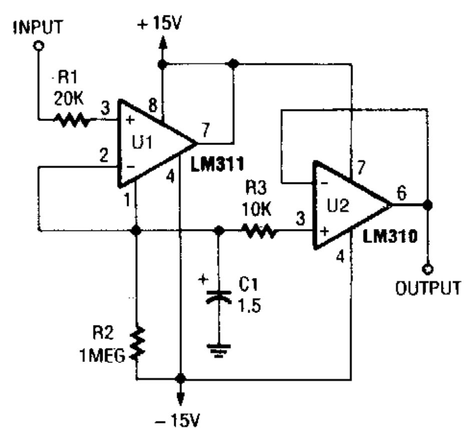
POPULAR ELECTRONICS *Fig. 26-16*

NEGATIVE PEAK DETECTOR



POPULAR ELECTRONICS *Fig. 26-17*

POSITIVE PEAK DETECTOR



POPULAR ELECTRONICS *Fig. 26-18*

455-kHz AM DETECTOR

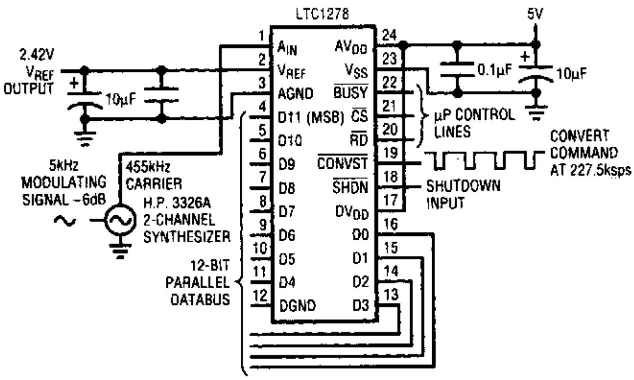


Figure A. The LTC1278 undersamples the 455-kHz carrier to recover the 5-kHz modulating signal

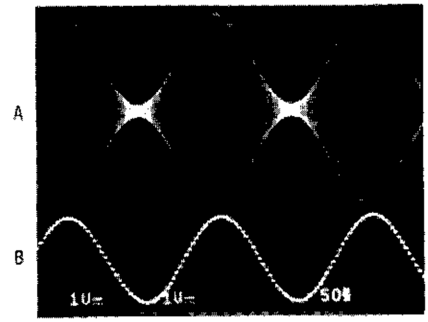
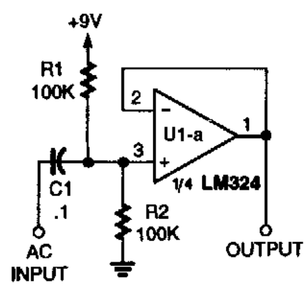


Figure B. Demodulating an IF by undersampling

LINEAR TECHNOLOGY *Fig. 26-19*

The LTC1278 undersamples the 455-kHz carrier to recover the 5-kHz modulating signal. The application shown uses the LTC1278 to undersample (at 227.5 ksps) a 455-kHz IF amplitude-modulated by a 5-kHz sine wave. Figures 26-19A and 26-19B show, respectively, the 455-kHz IF carrier and the recovered 5-kHz sine wave that results from a 12-bit DAC reconstruction.

ac NOISE DETECTOR



POPULAR ELECTRONICS *Fig. 26-20*

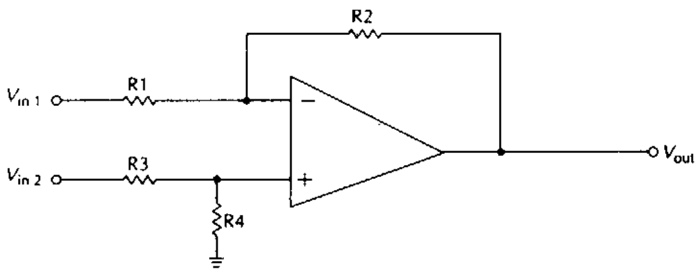
This circuit can be added to the glitch detector to trigger on ac noise.

27

Differential Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

BASIC OP-AMP DIFFERENTIAL AMPLIFIER



McGRAW-HILL

Fig. 27-1

In most cases, R_1 is equal to R_2, and R_3 has the same value as R_4. These equalities don't always have to be true, but they do significantly simplify the circuit design in most practical applications. In any case, for a true differential amplifier, the R_3:R_1 and R_4:R_2 ratios must be equal. That is:

$$\frac{R_3}{R_1} = \frac{R_4}{R_2}$$

The circuit still functions even if these ratios are not maintained, but the signals at the inverting and noninverting inputs are subjected to differing amounts of gain, which would be undesirable in most practical applications.

$$R_1 = R_2$$
$$R_3 = R_4$$

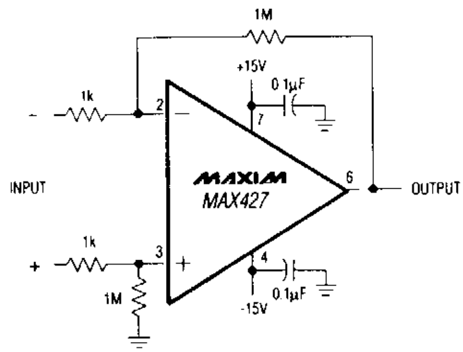
These resistance ratios determine the gain of the amplifier:

$$G = \frac{R_3}{R_1} = \frac{R_4}{R_2}$$

Assuming that the resistance ratios are maintained, the output voltage is equal to the differences between the two input voltages, multiplied by the gain. That is,

$$V_{OUT} = G \times (V_1 - V_2)$$

PRECISION HIGH-GAIN DIFFERENTIAL AMP



MAXIM *Fig. 27-2*

This circuit has a gain of 60 dB and a gain bandwidth of 8 MHz.

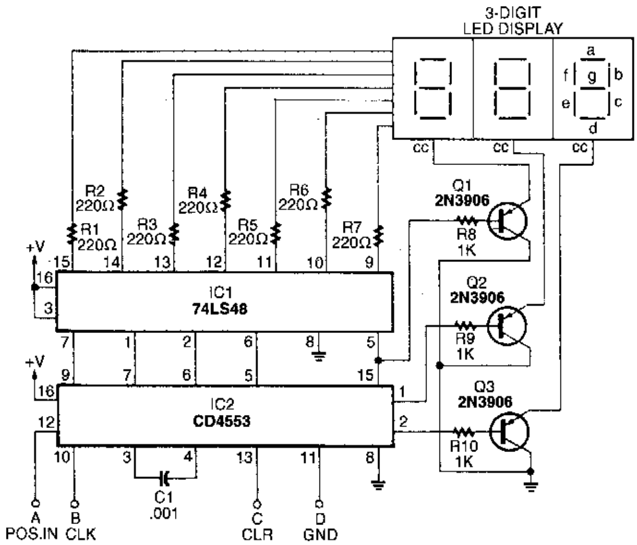
28

Display Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

Multiplexed BCD Decoder-Driver Circuit
Color-Shifting LED Display
Stereo Level Display
High-Efficiency Display Contrast and Backlight Control
Bar-Graph Level Gauge
Simple Color Organ
Voice Level Meter
LCD Contrast Temperature Compensator
LED Bargraph Driver Circuit

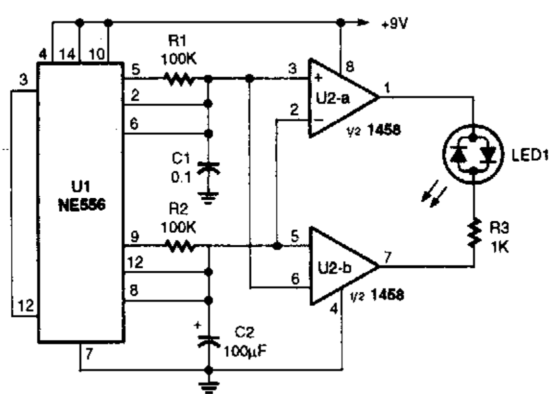
MULTIPLEXED BCD DECODER-DRIVER CIRCUIT



ELECTRONICS NOW *Fig. 28-1*

The BCD decoder-driver circuit will interface with any standard BCD output to produce a digital display.

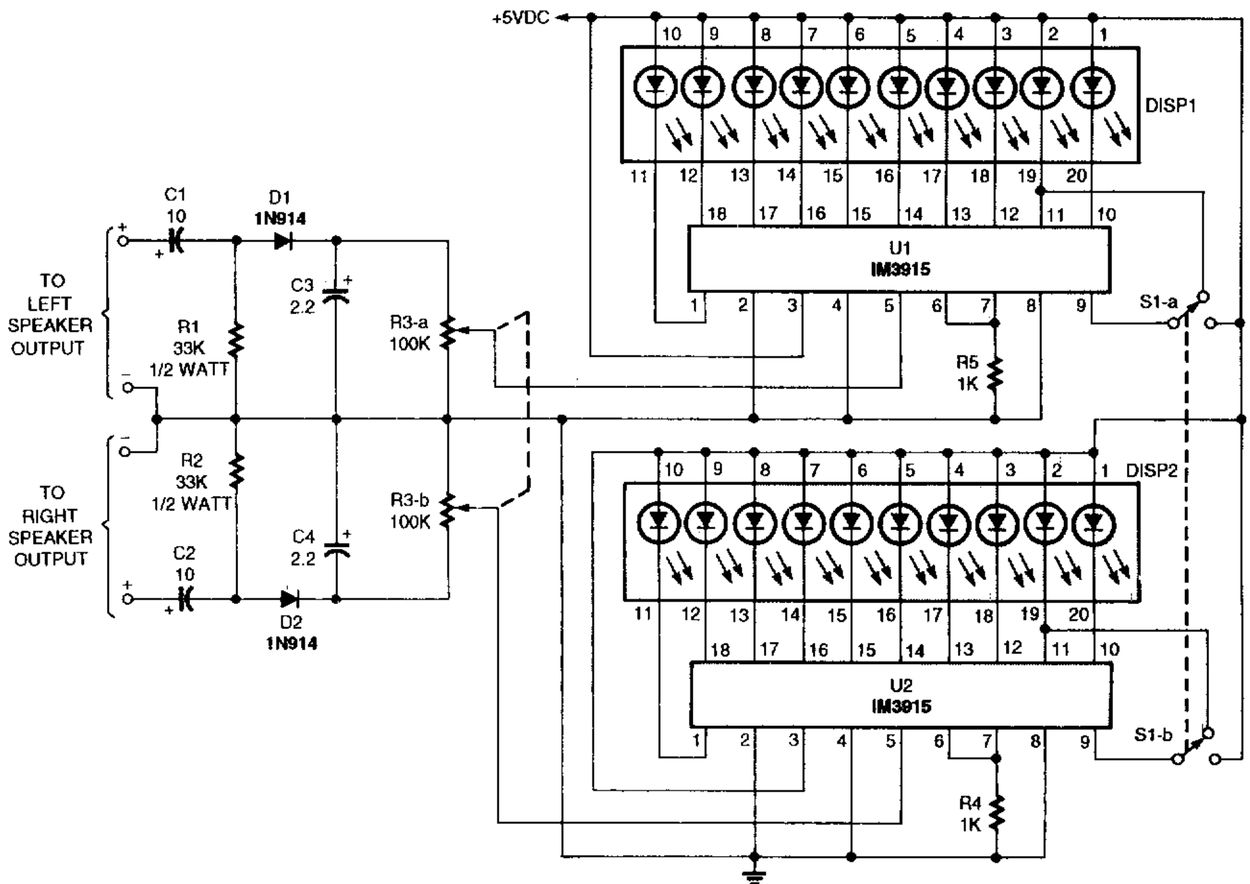
COLOR-SHIFTING LED DISPLAY



POPULAR ELECTRONICS *Fig. 28-2*

This circuit is used to make a tricolor LED gradually change color from yellow to red to yellow to green, and then back to yellow, where the cycle repeats. It is very simple to make, and the theory of operation is also simple. Both of the timers in the 556 dual oscillator/timer are configured for astable operation with a 50% duty cycle. One timer is set to oscillate much faster than the other. The timing capacitor voltage of each is sent to two comparators, which apply a voltage across the tri-color LED whose polarity depends on which capacitor voltage is higher. The rapidly changing capacitor's voltage causes the red and green elements of the LED to be alternately lit, thus giving the illusion of yellow light. As the slowly rising and falling voltage from the slower trimming capacitor changes in average value, it shifts the duty cycle to favor one color or the other. That gives the transition between colors a smooth appearance.

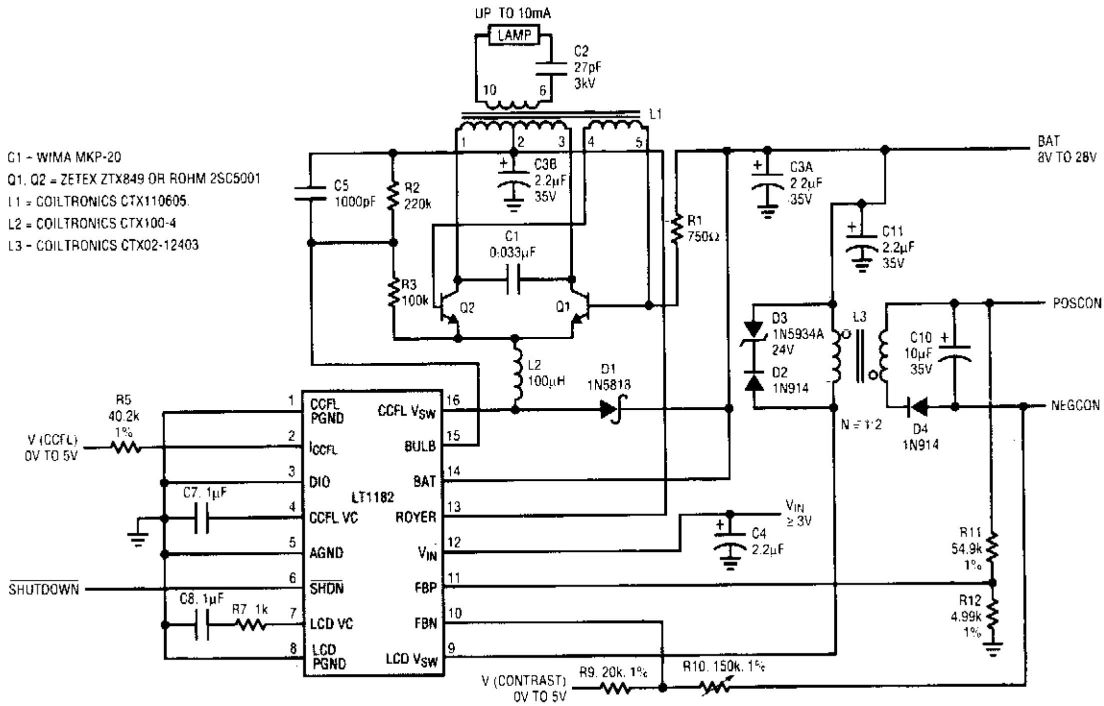
STEREO LEVEL DISPLAY



POPULAR ELECTRONICS *Fig. 28-3*

Two bar graph drivers and LEDs are used in this volume (level) indicator. R3a and R3b set the sensitivity of the circuit. The LEDs can be either bar graph units or individual LEDs can be used.

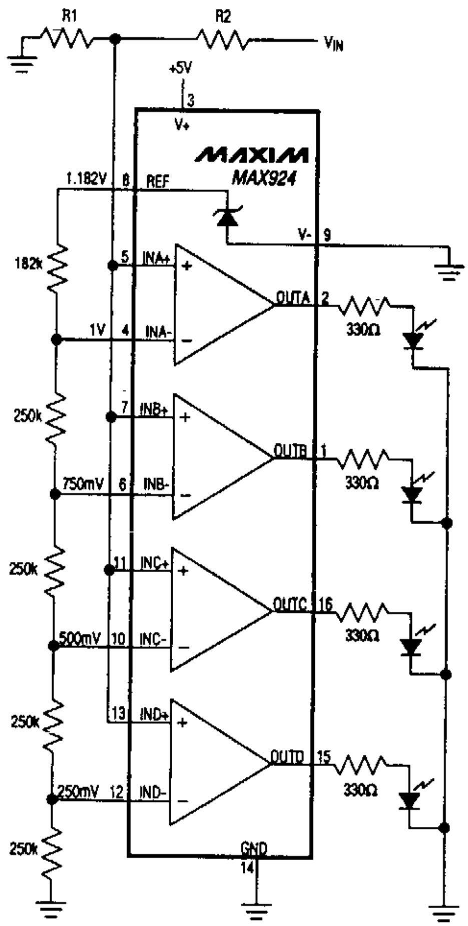
HIGH-EFFICIENCY DISPLAY CONTRAST AND BACKLIGHT CONTROL



LINEAR TECHNOLOGY *Fig. 28-4*

The LT1182 and LT1183 are compact high-performance solutions for powering LCD screens used in portable computers and instruments. Backlight control using a Cold Cathode Fluorescent lamp (CCFL) is accomplished with a switching regulator at efficiencies up to 90%. A second switching regulator converts the positive input to either positive or negative bias voltages used for LCD contrast control. Both regulators allow full range of adjustment using a D/A converter, PWM or potentiometer control. Grounded bulb configurations are also easily controlled with minimal parts count. The 200-kHz switching frequency minimizes the size of transformers and external components. A shutdown mode powers down both regulators and reduces supply current to just 35 μA.

BAR-GRAPH LEVEL GAUGE



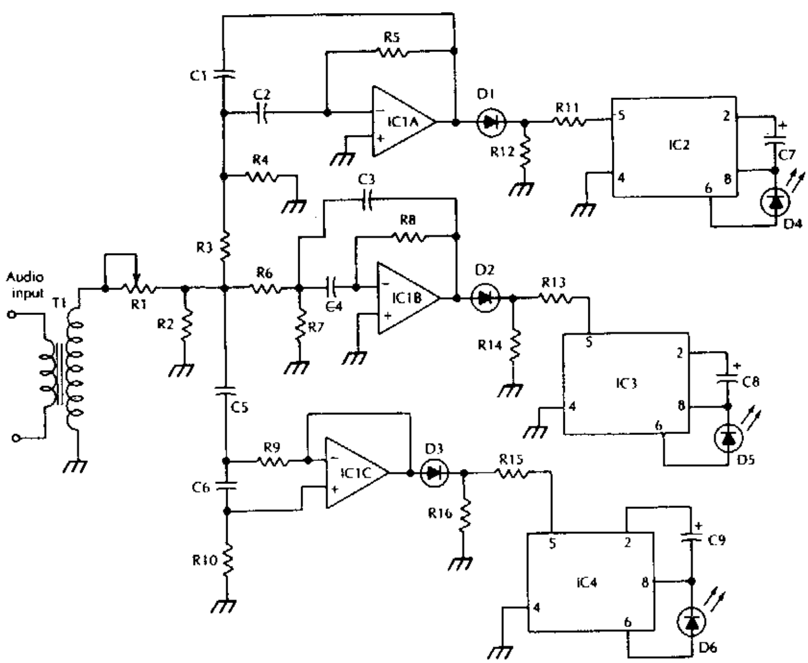
MAXIM

Fig. 28-5

A quad comparator and divider network form a 4-level comparator that drives 4 LEDs. R1 and R2 can be used to scale the basic sensitivity of 250 mV/LED to higher voltages than the basic 1 volt (for all 4 LEDs lit).

SIMPLE COLOR ORGAN

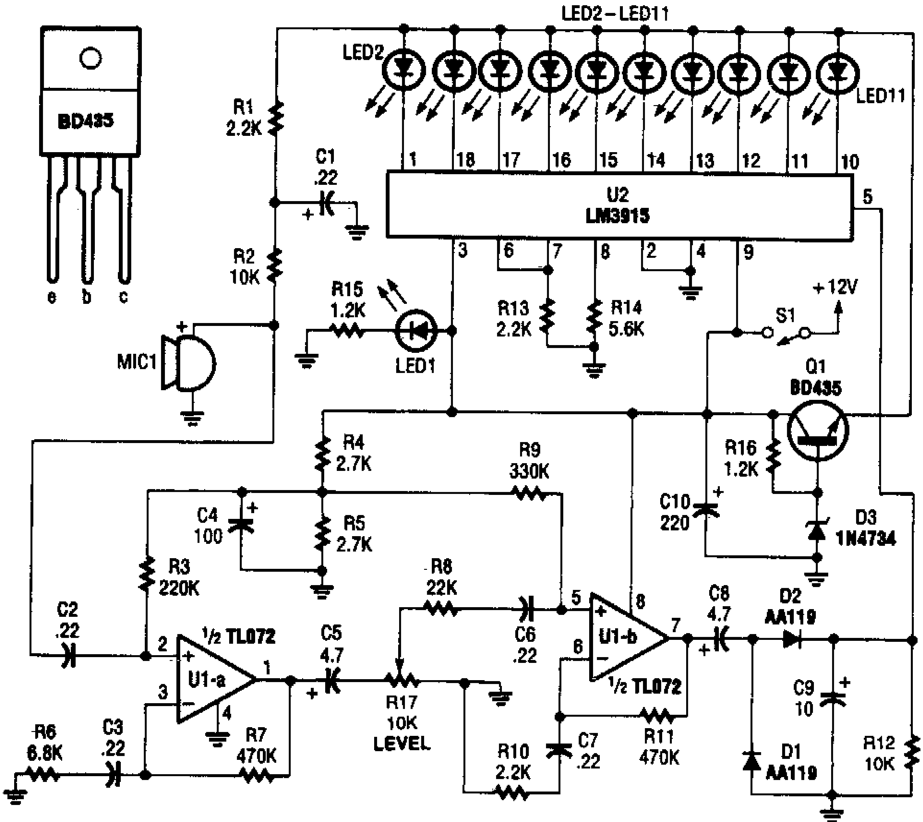
IC1	op amp
IC2, IC3, IC4	LM3909 LED flasher/oscillator
D1, D2, D3	diode (1N914, 1N4148, or similar)
D4, D5, D6	LED
T1	impedance-matching transformer (8 Ω:1 kΩ—see text)
C1, C2, C3, C4	0.1 μF capacitor
C5	0.047 μF capacitor
C6	0.01 μF capacitor
C7, C8, C9	47 μF 6 V electrolytic capacitor
R1	100 kΩ potentiometer
R2	47 kΩ ¼ W 5% resistor
R3	2.2 kΩ ¼ W 5% resistor
R4	680 Ω ¼ W 5% resistor
R5	220 kΩ ¼ W 5% resistor
R6	390 Ω ¼ W 5% resistor
R7	1.2 kΩ ¼ W 5% resistor
R8, R10, R12, R14, R16	10 kΩ ¼ W 5% resistor
R9	3.3 kΩ ¼ W 5% resistor
R11, R13, R15	33 kΩ ¼ W 5% resistor



McGRAW-HILL *Fig. 28-6*

Three active filters that divide the audio spectrum into three bands drive rectifiers and then drive IC2, 3, and 4, flashing the LEDs at 6 Hz. D4, D5, and D6 should be three different colors for best effect.

VOICE LEVEL METER

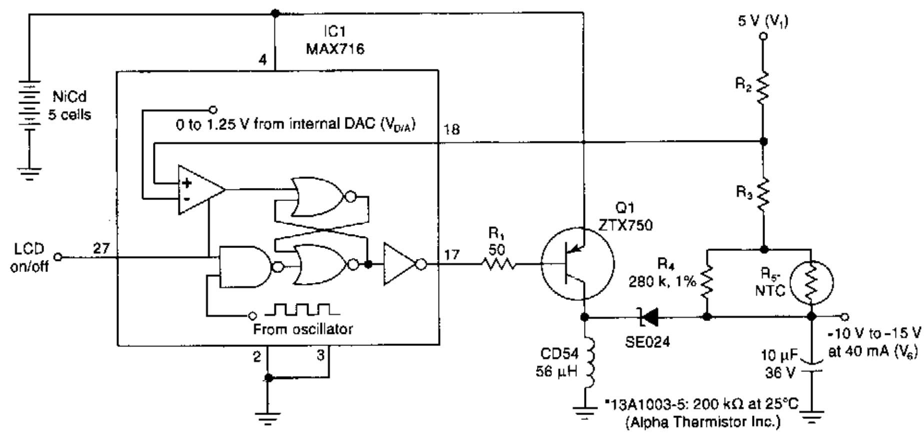


This volume meter can be handy anywhere you need to measure the relative sound level in a room. It is readily adjustable to increase its usefulness.

POPULAR ELECTRONICS *Fig. 28-7*

Using an LM3915 VU meter LED bar graph driver, 10 LEDs are driven. A simple audio amplifier drives a detector circuit, which provides dc drive for the LM3915.

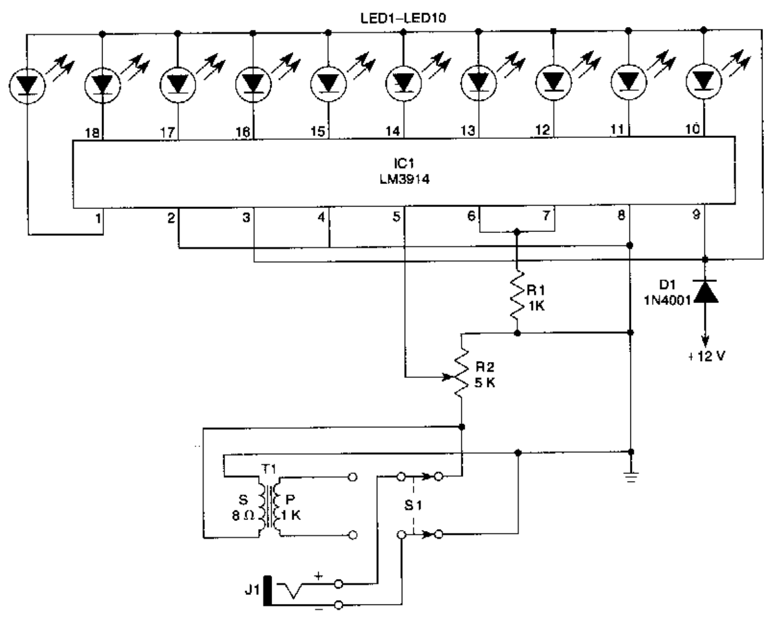
LCD CONTRAST TEMPERATURE COMPENSATOR



ELECTRONIC DESIGN *Fig. 28-8*

Negative temperature-coefficient resistor R5 modifies feedback in this switching regulator, which results in a negative output voltage that varies with temperature. With properly chosen resistor values, the circuit produces a temperature-compensated bias voltage that ensures constant contrast in an LCD.

LED BARGRAPH DRIVER CIRCUIT



ELECTRONICS NOW *Fig. 28-9*

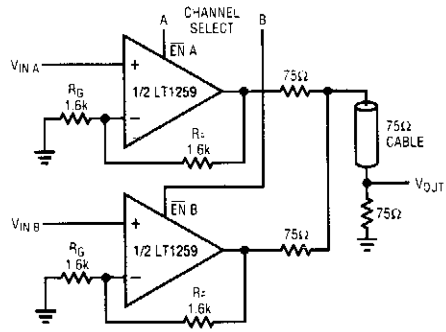
This circuit is used as an audio indicator. S1 selects direct input or a 1-kΩ (high impedance) audio input. R2 is a sensitivity control.

29

Driver Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

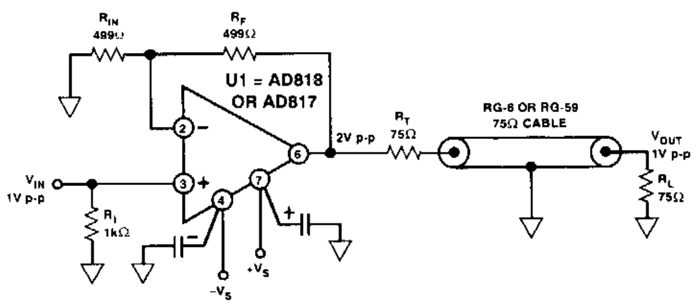
TWO-INPUT VIDEO MUX CABLE DRIVER



LINEAR TECHNOLOGY

Fig. 29-1

IMPEDANCE-MATCHED LINE DRIVER WITH 75-Ω LOAD

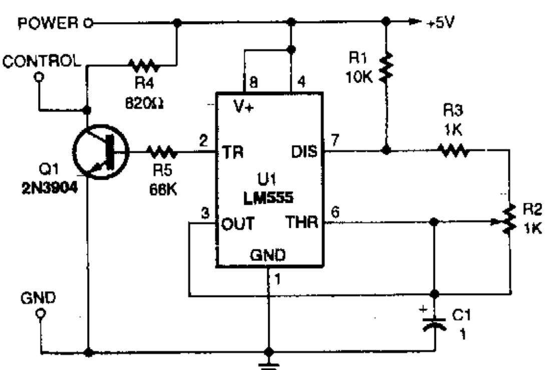


ANALOG DIALOG

Fig. 29-2

This circuit is a wideband 75-Ω line driver, for video applications (1 V p-p into 75 Ω).

TESTS DRIVER FOR HOBBY SERVOS

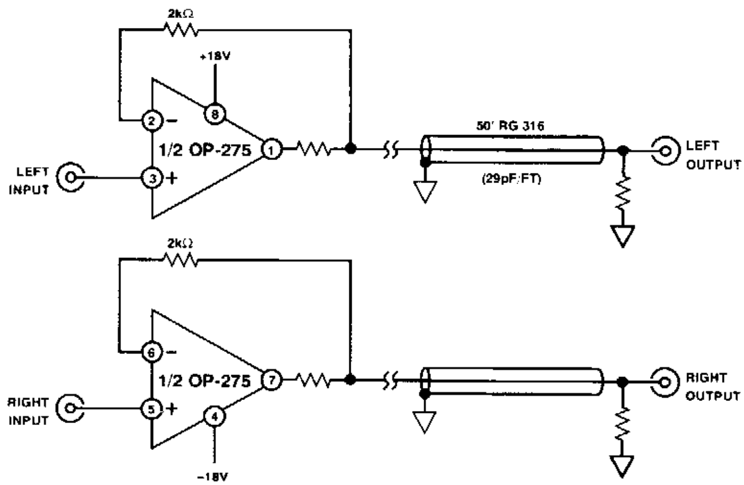


This circuit will generate the pulse used to control hobby servos. With the components shown the servo should produce a 90° total rotation.

POPULAR ELECTRONICS

Fig. 29-3

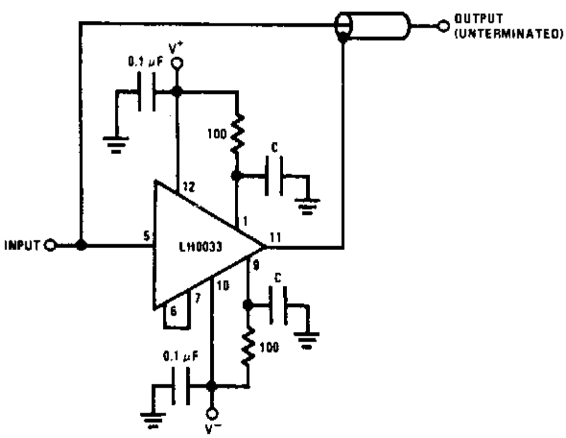
STEREO LINE DRIVER



ANALOG DEVICES ***Fig. 29-4***

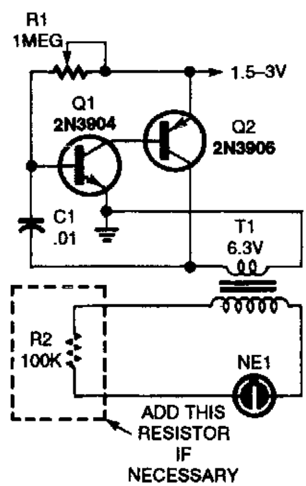
One Analog Devices OP-275 can be used for stereo line driver applications.

HIGH-SPEED SHIELD/LINE DRIVER



NATIONAL SEMICONDUCTOR ***Fig. 29-5***

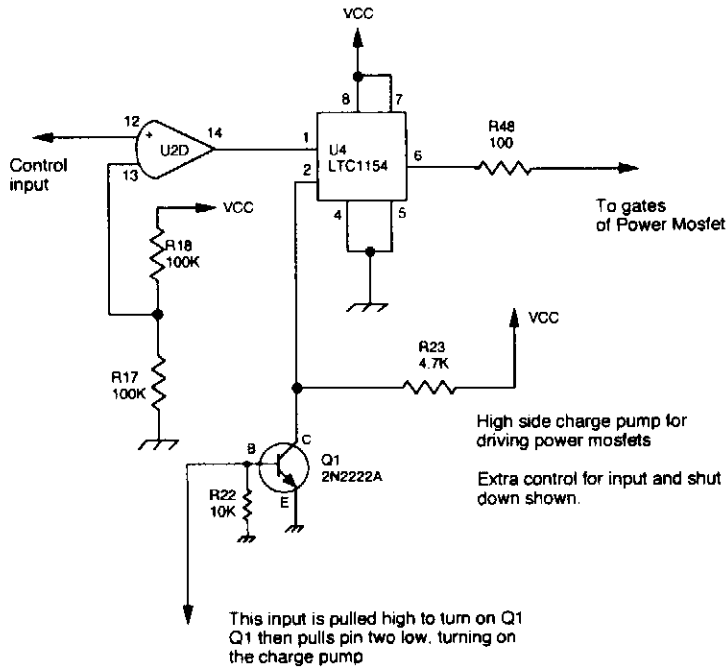
SIMPLE NEON LIGHT DRIVER



POPULAR ELECTRONICS ***Fig. 29-6***

NE1, a neon lamp, is lit by this simple inverter circuit. T1 is a 20:1 turn ratio transformer (transistor radio output, etc.).

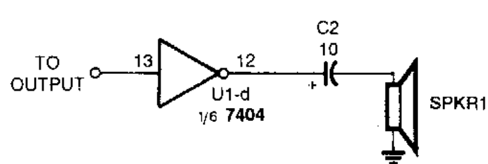
HIGH-SIDE MOSFET DRIVER



73 AMATEUR RADIO TODAY ***Fig. 29-7***

A Linear Technology LTC1154 is used as a charge pump to drive the gate of a high-side power MOSFET.

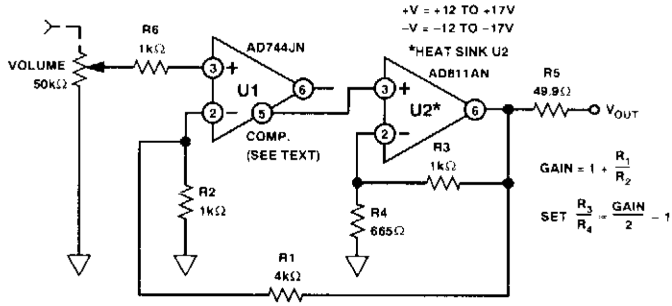
TTL-BASED SPEAKER DRIVER



POPULAR ELECTRONICS ***Fig. 29-8***

A TTL IC, such as a 7404, can drive a small speaker with enough audio to be used as an alarm or annunciator. The speaker can be a 32- or 100-Ω unit.

LOW-DISTORTION COMPOSITE ± 100-mA LINE DRIVER

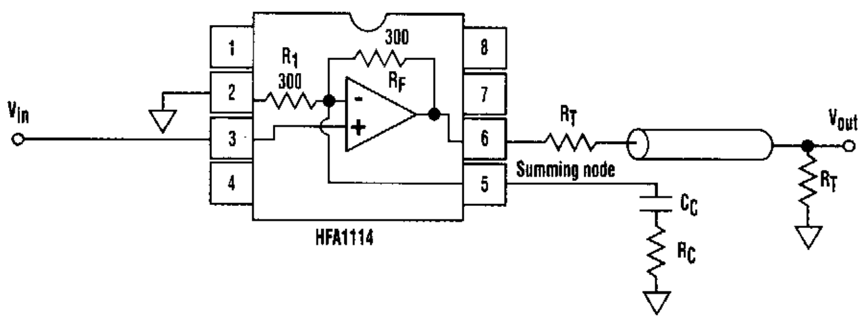


ANALOG DEVICES

Fig. 29-9

This line driver combines the high input impedance of an FET-input IC and a 100-mA op amp. U1's output is left open. The compensation terminal (pins) drive U2's high-*Z* input for increased overall phase margin. Gain is 14 dB, THD +*N* at 5 V, and RMS output is around 0.001% below 20 kHz.

VIDEO CABLE DRIVER



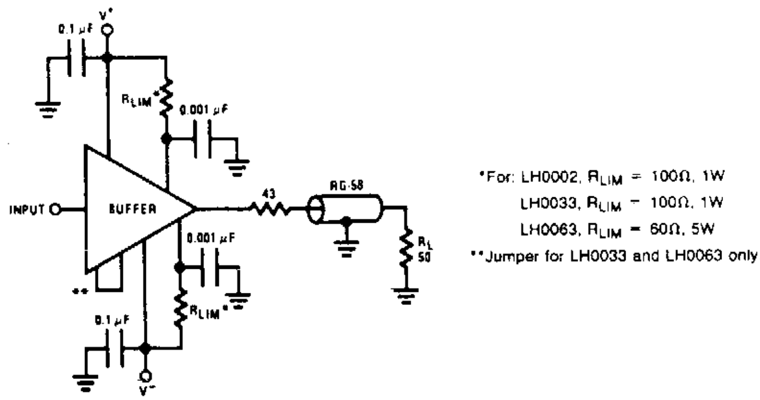
ELECTRONIC DESIGN

Fig. 29-10

The driver's frequency response is tunable for a specific cable length via components connected to the summing node. By shunting R1, R_c acts to increase the amplifier's gain, and C_c controls the cut-in frequency of the compensation.

These three components peak the amplifier's frequency response to counteract the cable's roll-off characteristic. By squeezing more bandwidth out of a given cable, higher-performance cables aren't needed.

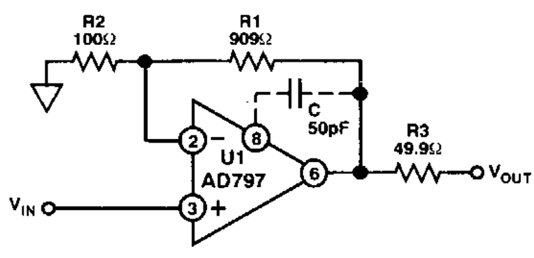
COAX CABLE DRIVER



NATIONAL SEMICONDUCTOR *Fig. 29-11*

Because of their high-current drive capability, the LH0002, LH0033, and LH0063 buffer amplifiers are suitable for driving terminated or unterminated coaxial cables, and high-current or reactive loads. Current-limiting resistors should be used to protect the device from excessive peak load currents or accidental short circuit. No current limiting is built into the devices other than that imposed by the limited beta of the output transistors. This figure shows a coaxial-cable drive circuit. The 43-Ω resistor is included, the output voltage to the load is about half what it would be without the near-end termination.

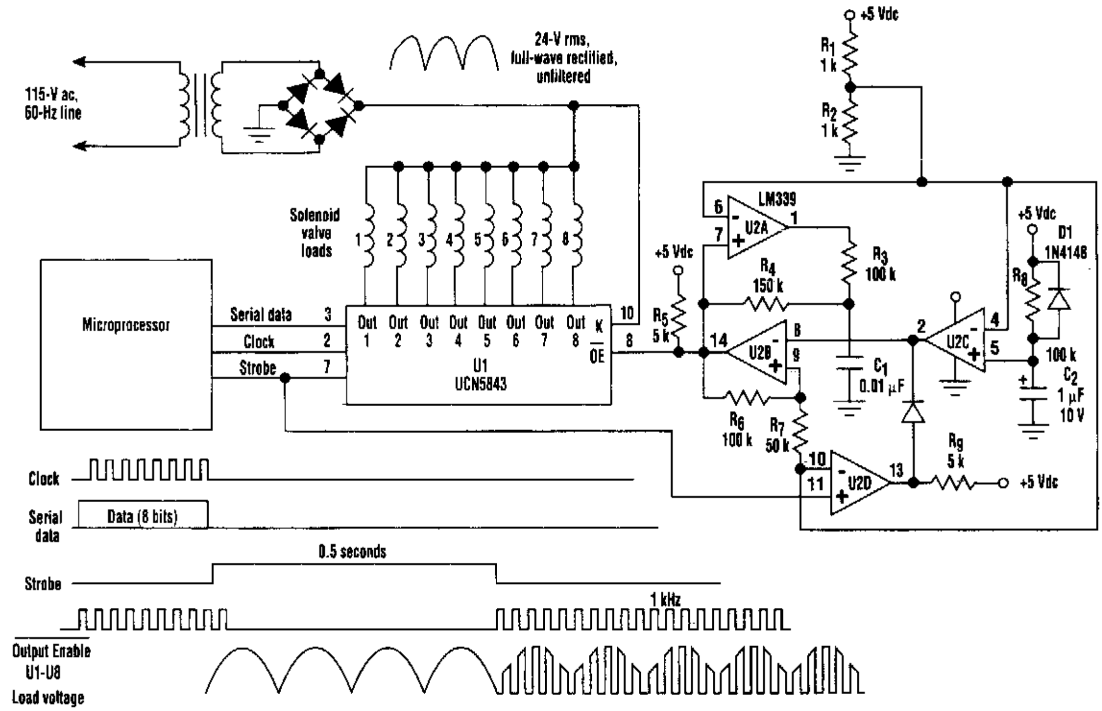
ULTRA LOW DISTORTION ±50-mA DRIVER



ANALOG DEVICES *Fig. 29-12*

For a 600-Ω load, THD is typically −115 dB at 20 kHz, 3-V RMS output, with ±15-V supplies. The −3-dB BW is 6 MHz.

VERY EFFICIENT SOLENOID DRIVER



ELECTRONIC DESIGN *Fig. 29-13*

In the circuit shown, the IC in the center, U1, contains a microprocessor-compatible serial-in, parallel-out shift register, with strobed latch and high-level solenoid drives (see the figure).

The strobe pulse turns on the selected loads at full power for its pulse width, which is 0.5 second. Following the strobe pulse, the driver outputs are pulse-width modulated by the multivibrator, which is set for a duty cycle of 25%. The solenoids are therefore held in the engaged position with a voltage that's 25% of the nominal supply voltage.

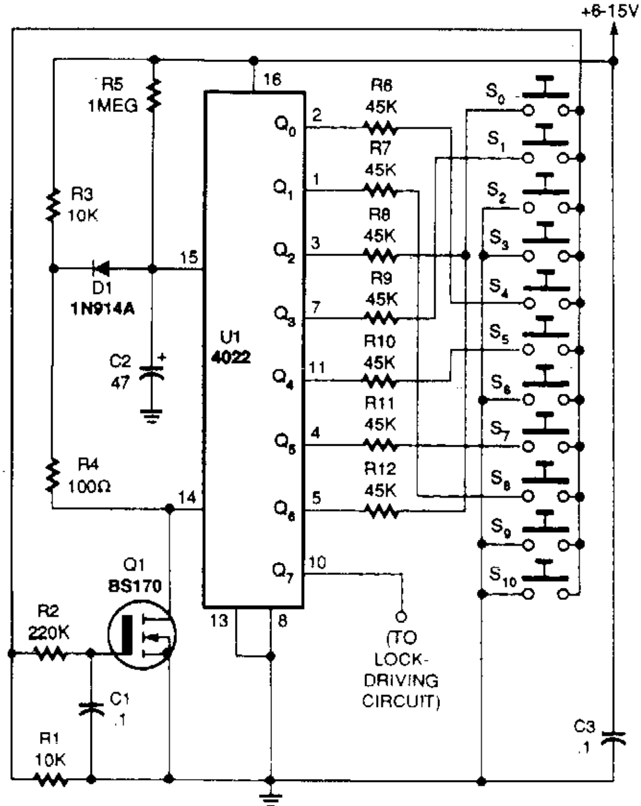
U2B acts as the basic multivibrator in the circuit, and U2A sets the duty cycle by setting the discharge current of capacitor C1. U2D overrides the oscillation during the strobe pulse. And U2C provides the Power-On Reset (POR), inhibiting the solenoid loads from turning on during the initialization period of the microprocessor.

30

Electronic Lock Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

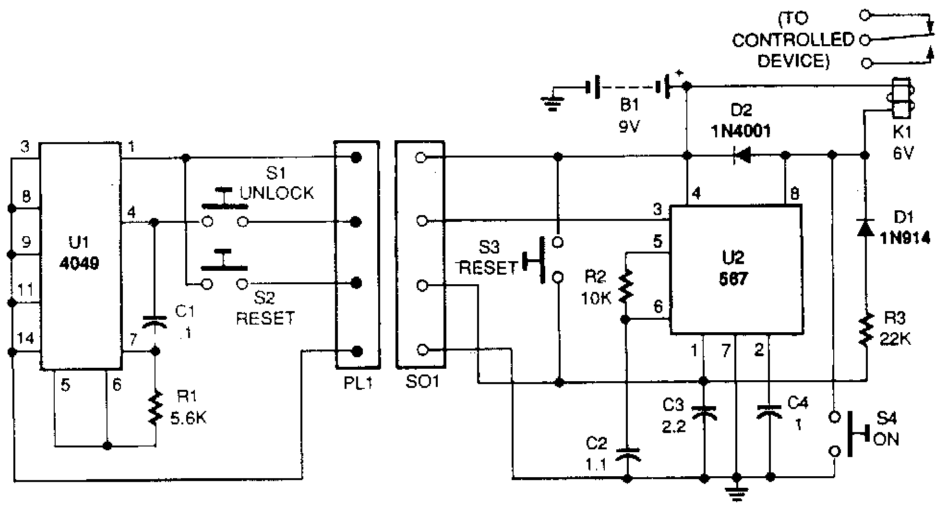
ELECTRONIC LOCK



POPULAR ELECTRONICS ***Fig. 30-1***

The heart of the circuit is a 4022 octal counter. When first powered up, C2 is charged via R5, so the reset input of the counter is kept high. That causes output Q_0 to go high while all other outputs are low. With the switches wired as shown, when S4 is pressed, the BS170 is switched on via debouncing network R2/C1, and U1 receives a clock pulse. Also, C2 is discharged via R4 and D1, removing the reset signal of the counter, allowing it to advance. The time required for C2 to charge via R5 (i.e., to reset the counter), is the maximum time that can lapse before the next key is pressed. The above cycle is therefore repeated only if S8 (connected to the Q1 output) is pressed in time. When all keys have been pressed in time and in the correct order, Q7 goes high for about four seconds to drive the "unlock" circuitry (e.g., a relay driver for an automatic door opener. A builder can change the code by reviewing the switches. The code for the lock shown in the circuit diagram is 4-8-0-1-5-7-0. However, the 4022 octal counter can be replaced by a 4017 divide-by-10 counter. That will make it possible to add two more digits to the combination.

FREQUENCY-BASED LOCK



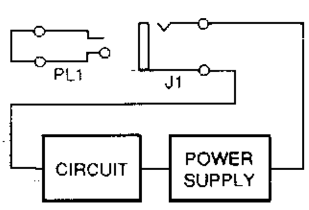
POPULAR ELECTRONICS ***Fig. 30-2***

The system is formed by two separate circuits—a key and a keyhole. The key engages the keyhole with a mating pair of connectors. The key is a tone-generator circuit consisting of a 4049 hex inverter CMOS IC (U1), switches (S1 and S2), a resistor (R1), and a capacitor (C1). The value of the tone generated by that circuit in Hz is determined by;

$$\frac{1}{(1.4 \times R_1 C_1)}$$

The keyhole is a 567 tone-decoder circuit that can be configured to detect any frequency from 0.01 Hz to 500 kHz. The frequency it detects (f_0), via the 567 IC, turns on the relay (K1). Components R3 and D1 are used to latch the circuit, so the output stays on even after the input tone is removed. When S2 is pressed, the system is reset. Switch S3 resets the circuit from inside.

SIMPLE LOCK



POPULAR ELECTRONICS ***Fig. 30-3***

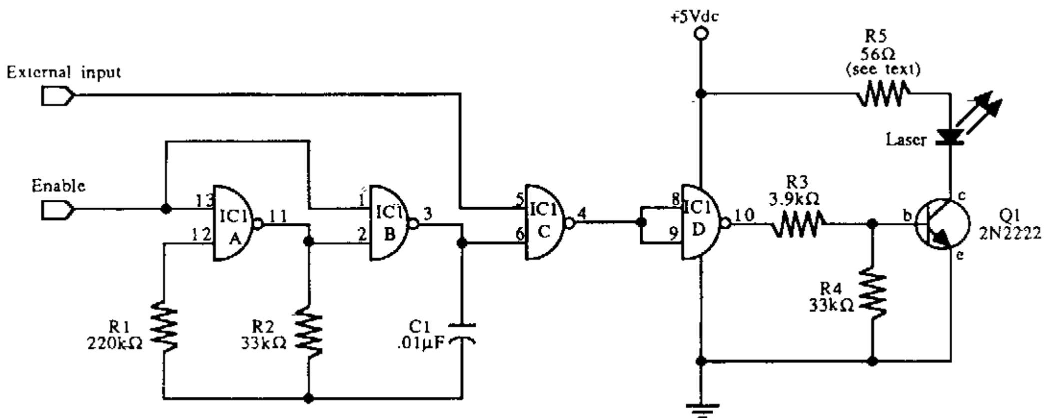
Only an appropriately wired plug of the right size will activate circuits with a nonshorting jack in their power supply circuit.

31

Fiber-Optics Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

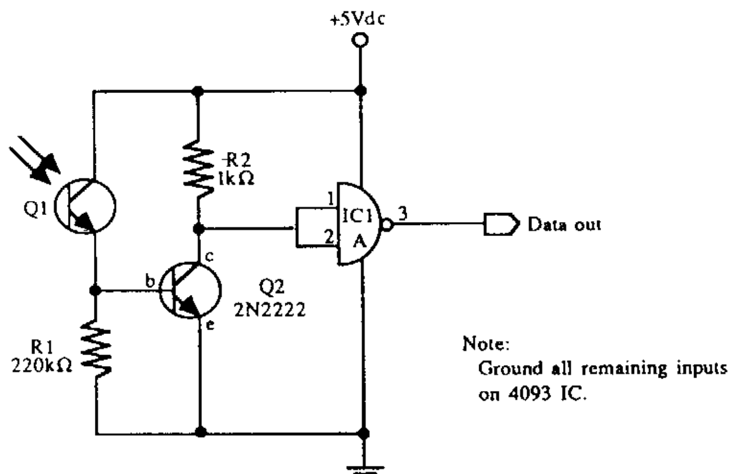
EXPERIMENTAL DATA TRANSMITTER FOR FIBER OPTICS



McGRAW-HILL *Fig. 31-1*

This schematic for an experimental data transmitter uses optical fibers and a laser diode. Transmission frequency of the free-running oscillator is approximately 3 kHz. R5 might have to be varied to suit your laser diode. IC1 is a CD4093.

EXPERIMENTAL FIBER-OPTIC DATA RECEIVER



McGRAW-HILL *Fig. 31-2*

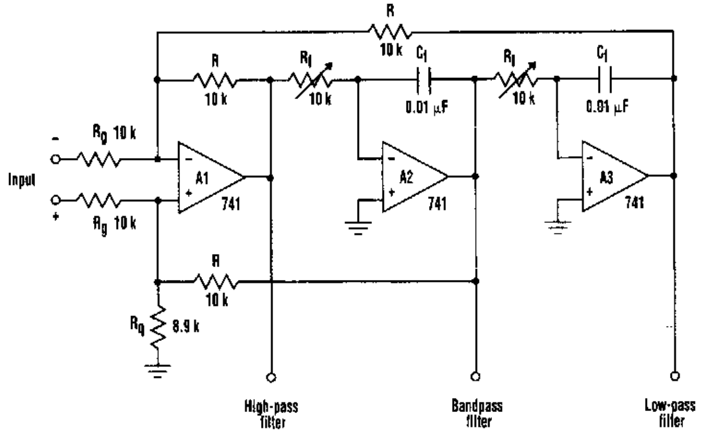
An infrared phototransistor acts as the sensor for this receiver. IC1a is a section of a CD4093 CMOS NAND gate.

32

Filter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

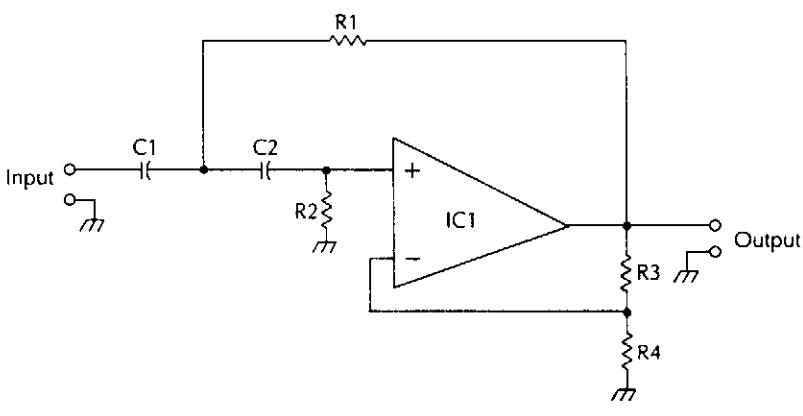
STATE VARIABLE FILTER



ELECTRONIC DESIGN *Fig. 32-1*

The state variable filter shown consists of only three op amps and a few passive components. It provides several key features. These include the ability to simultaneously provide low-pass, high-pass, and bandpass filter functions, and adjust bandwidth in a wide range by changing the values of C_f and R_f. The device also is easy to tune and simple to construct, while the quality factor (Q) of each filter is independent of each other.

SALLEN-KEY HIGH-PASS FILTER



McGRAW-HILL *Fig. 32-2*

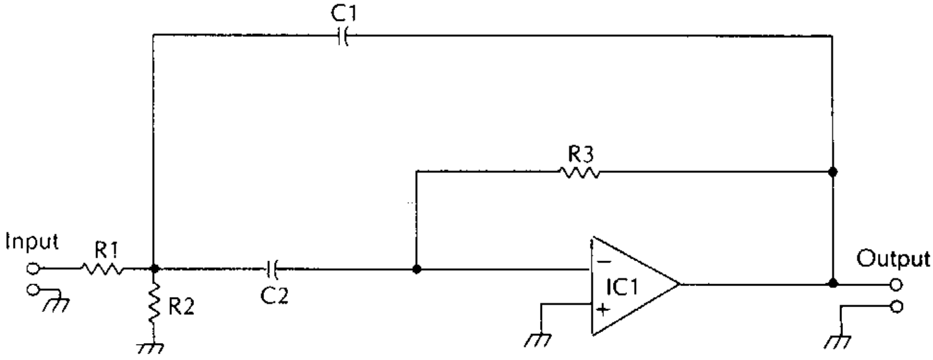
R3 and R4 set the circuit gain

$$f_c = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

usually $C_1 = C_2, R_1 = R_2$

$$R_3 = 0.586 R_4$$

ACTIVE BANDPASS FILTER CIRCUIT



MCGRAW-HILL *Fig. 32-3*

In this circuit,

$$C_1 = C_2 = C$$

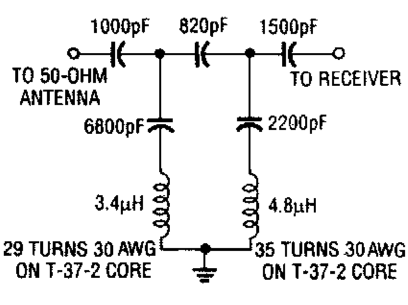
$$Q = \frac{f_0}{BW} \quad K = circuit\ gain,\ f_0 = center\ frequency$$

$$R_1 = \frac{Q}{2\pi f_0 CK}$$

$$R_2 = \frac{2Q}{2\pi f_0 C}$$

$$R_3 = \frac{Q}{2\pi f_0 C(2Q - K)}$$

HIGH-PASS FILTER

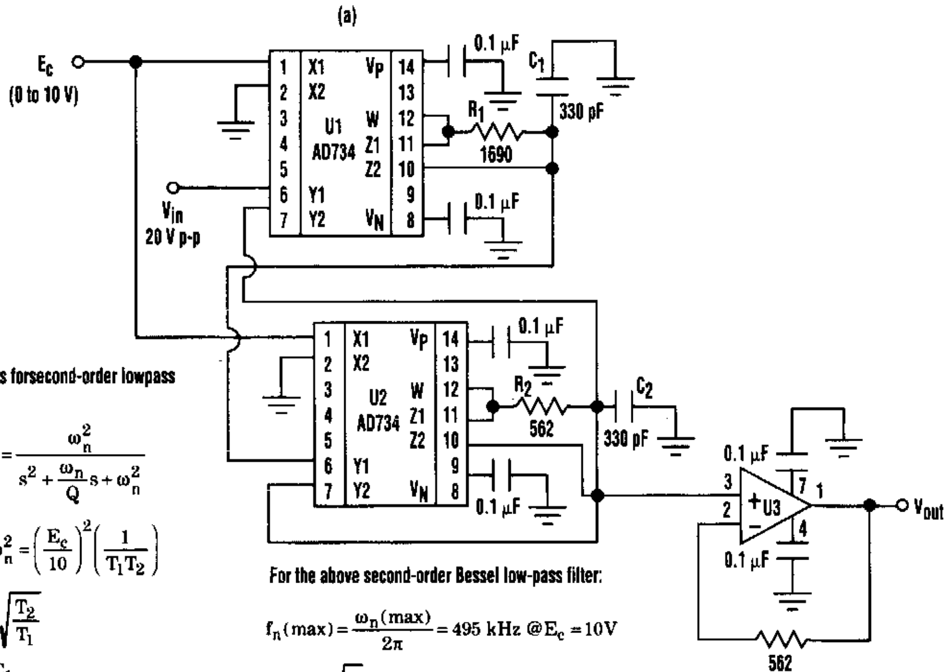


RADIO-ELECTRONICS *Fig. 32-4*

This high-pass filter will attenuate AM stations by 40 dB. Its low-frequency cutoff is about 2.2 MHz. This filter is useful for SW listening in areas of high AM radio signal strength.

SECOND-ORDER VOLTAGE-CONTROLLED FILTER

(a)



Design equations forsecond-order lowpass filter:

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{\omega_n^2}{s^2 + \frac{\omega_n}{Q}s + \omega_n^2}$$

$$\text{where } \omega_n^2 = \left(\frac{E_c}{10}\right)^2 \left(\frac{1}{T_1 T_2}\right)$$

$$\text{and } Q = \sqrt{\frac{T_2}{T_1}}$$

$$T_1 = R_1 C_1$$

$$T_2 = R_2 C_2$$

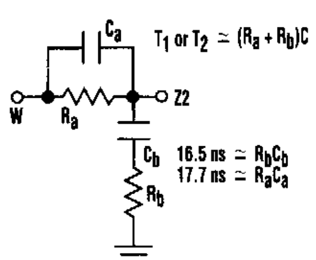
For the above second-order Bessel low-pass filter:

$$f_n(\max) = \frac{\omega_n(\max)}{2\pi} = 495 \text{ kHz } @E_c = 10V$$

$$\text{and } Q = \sqrt{\frac{1}{3}}$$

(b)

Compensation network if T_1 or T_2 < 200 ns

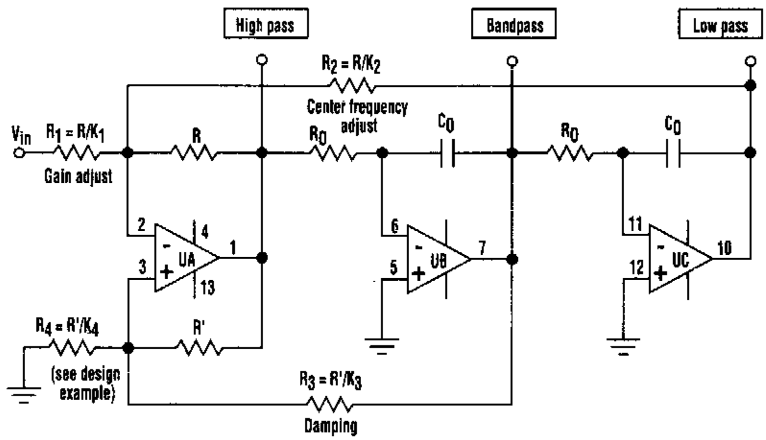


ELECTRONIC DESIGN

Fig. 32-5

Desirable second-order voltage-controlled low-pass filter response can be achieved with this voltage-controlled filter (A). By using low-distortion, wide-bandwidth multipliers, it achieves higher cutoff frequencies than switched-capacitor filters. If the circuit's RC network has a time constant less than 200 ns, it should be replaced by a lag compensator network (B).

COMBINATION FILTER



ELECTRONIC DESIGN

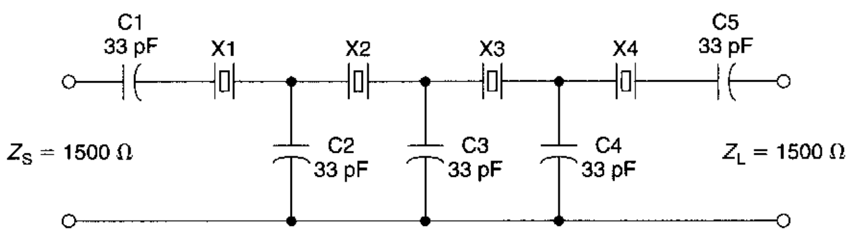
Fig. 32-6

The classic "state variable" two-integrator filter is known for its insensitivity to component variations, and its ability to provide three separate simultaneous outputs—low pass, high pass, and bandpass.

Typically, a quad op amp is used to implement the state-variable filter. The classic configuration uses two integrating amplifiers, a filter input amplifier, and a filter feedback amplifier.

The design described here combines both input and feedback amplifiers into one adder/subtractor amplifier, achieving a three op-amp filter design (see the figure).

SHORTWAVE RECEIVER IF FILTER



3-MHz IF filter
BW = 700 Hz
IL ≈ 4 dB

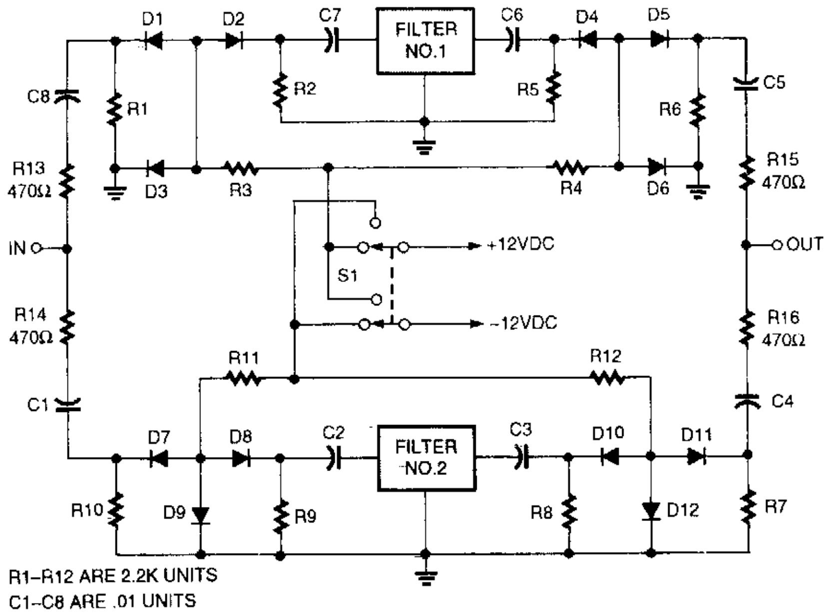
X1 through X4 3.000 MHz ±50 Hz
C_X = C1 through C5 33 pF ±5% NPO

WILLIAM SHEETS

Fig. 32-7

An inexpensive filter can be made from microprocessor crystals. This filter has 700 Hz BW (3 dB) and has a flat response (<1 dB) for about 400 to 500 Hz. Although a 3-MHz crystal was used, any frequency from 2 to 15 MHz (using fundamental crystal) should work, with appropriate scaling of components. Crystal resonant frequencies should match within 20% and preferably 10% of expected bandwidth (which is narrower as C_x increases. Impedance is reduced with wider bandwidths.

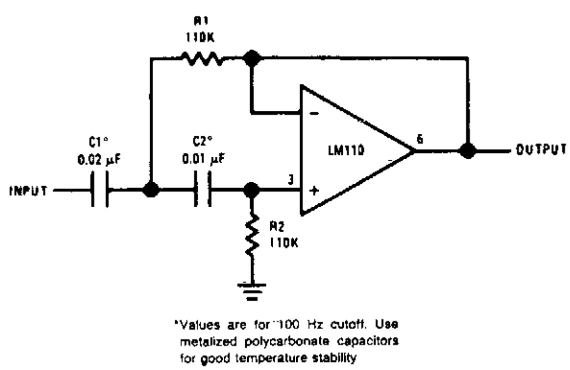
PIN DIODE FILTER SELECTION CIRCUIT



POPULAR ELECTRONICS ***Fig. 32-8***

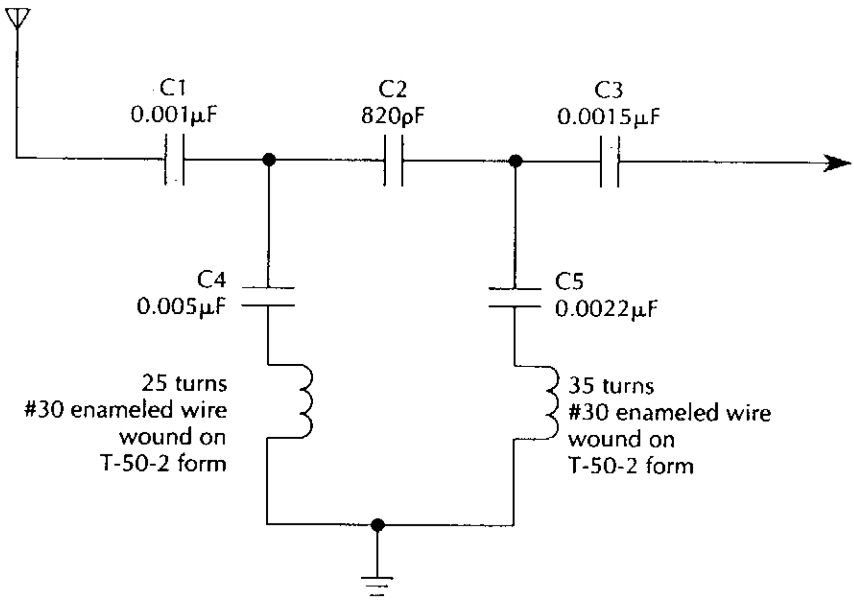
Selecting IF bandpass filters via series/shunt PIN-diode switching can be accomplished with this circuit.

HIGH-PASS ACTIVE FILTER



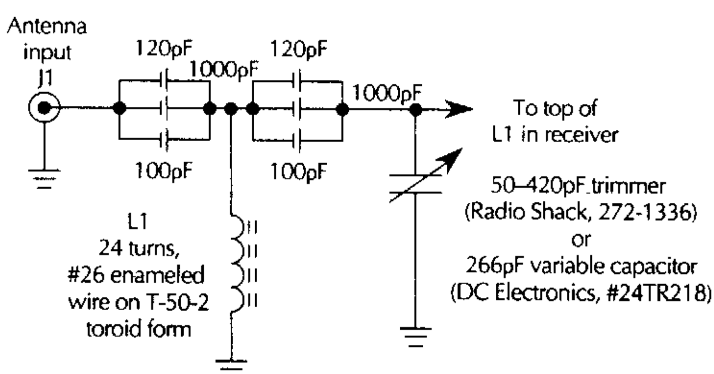
NATIONAL SEMICONDUCTOR ***Fig. 32-9***

AM BROADCAST TRAP FOR SIMPLE SW RECEIVERS



McGRAW-HILL *Fig. 32-10*

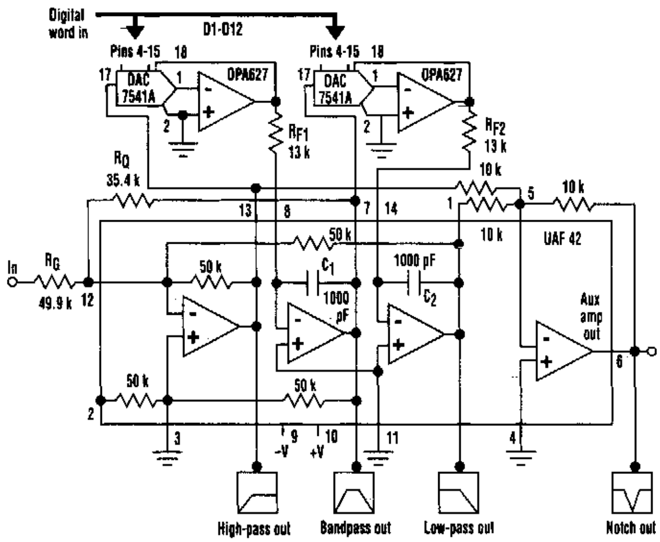
SHORTWAVE INTERFERENCE TRAP



McGRAW-HILL *Fig. 32-11*

Build this interference trap to help block strong shortwave, broadcast, and FM stations from coming in on the shortwave bands.

PROGRAMMABLE ANALOG FILTER

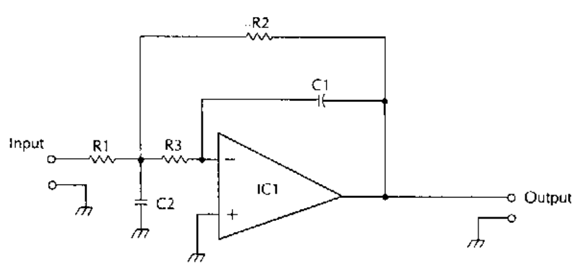


ELECTRONIC DESIGN

Fig. 32-12

The circuit in the figure shows how an analog, digitally programmable filter can be built using a UAF42. This monolithic, state-variable active filter chip provides a two-pole filter building block with low sensitivity to external component variations. It eliminates aliasing errors and clock feed though noise common to switched-capacitor filters. Low-pass, high-pass, bandpass, and notch (band-reject) outputs are available.

ACTIVE LOW-PASS FILTER



McGRAW-HILL

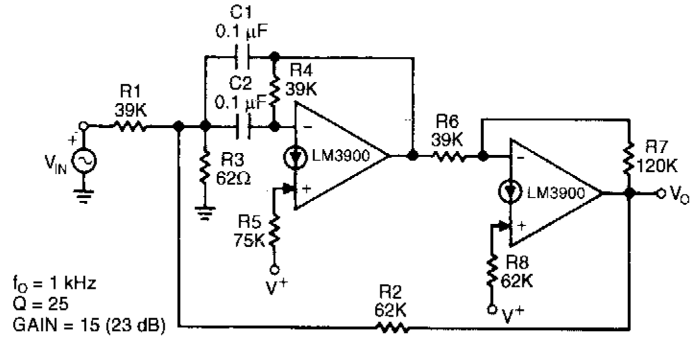
Fig. 32-13

In this circuit, $R_1 = 2$

$$R_2 = R_4$$
$$R_3 = 2R_1$$
$$C_1 = C_2$$
$$f = \frac{1}{2\pi RC}$$

This circuit has a rolloff of 6 dB/Octave.

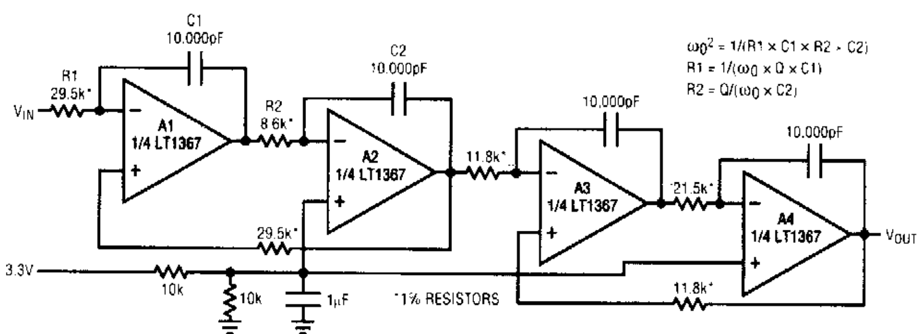
TWO OP-AMP BANDPASS FILTER



NATIONAL SEMICONDUCTOR *Fig. 32-14*

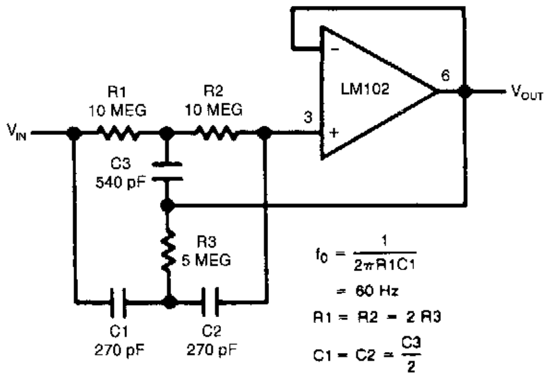
This circuit uses only two capacitors. The amplifier on the right supplies a controlled amount of positive feedback for improved response characteristics. Resistors R5 and R8 are used to bias the outvoltage of the amplifiers at $V+/2$.

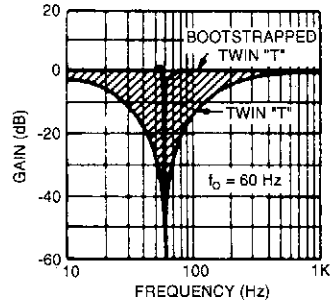
SINGLE 3.3-V SUPPLY 4-POLE STATE VARIABLE FILTER



LINEAR TECHNOLOGY *Fig. 32-15*

HIGH-Q NOTCH FILTER





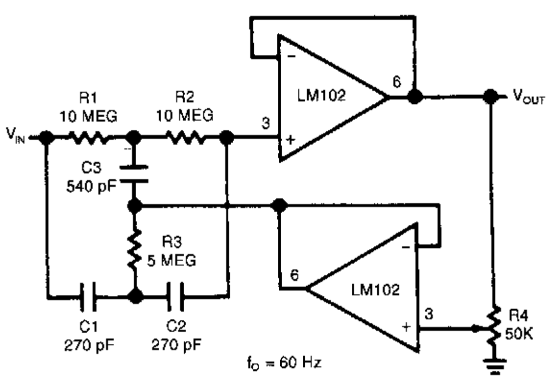
Response of High and Low Q Notch Filter

NATIONAL SEMICONDUCTOR

Fig. 32-16

This shows a twin "T" network connected to an LM102 to form a high Q, 60-Hz notch filter. The junction of R3 and C3, which is normally connected to ground, is bootstrapped to the output of the follower. Because the output of the follower is a very low impedance, neither the depth nor the frequency of the notch change; however, the Q is raised in proportion to the amount of signal fed back to R3 and C3.

ADJUSTABLE-Q NOTCH FILTER

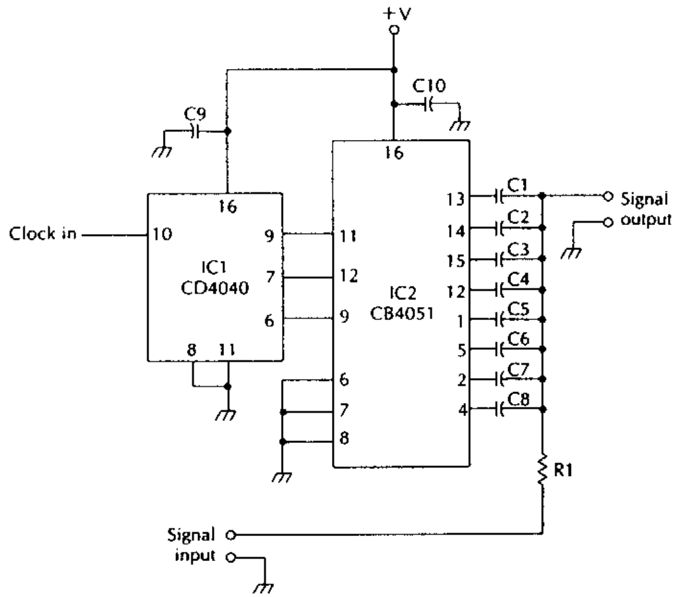


NATIONAL SEMICONDUCTOR

Fig. 32-17

This figure shows a circuit where the Q can be varied from 0.3 to 50. A fraction of the output is fed back to R3 and C3 by a second voltage follower, and the notch Q is dependent on the amount of signal fed back. A second follower is necessary to drive the twin "T" from a low-resistance source so that the notch frequency and depth will not change with the potentiometer setting.

DIGITAL COMB FILTER



This circuit uses an eight-step switching sequence, so n=8. This makes the center frequency equal to:

$$F_c = \frac{1}{(2nRC)}$$

$$= \frac{1}{(2 \times 8)RC}$$

$$= \frac{1}{16RC}$$

Using the component values suggested in the parts list, the circuit has a main center frequency of:

$$F_c = \frac{1}{(16 \times 1000 \times 0.00000001)}$$

$$= \frac{1}{0.00016}$$

$$= 6250 \text{ Hz}$$

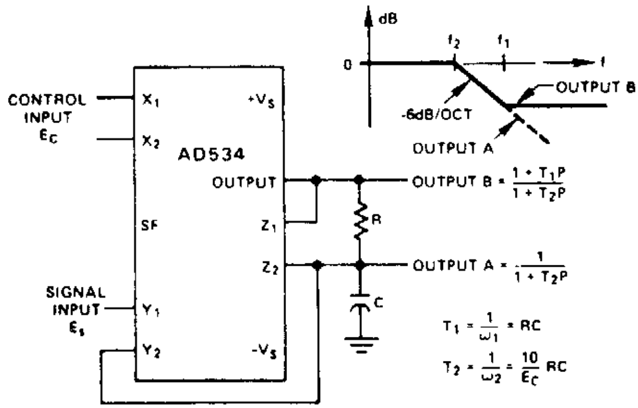
Suggested parts list for the digital comb filter

IC1	CD4040 BCD-ripple counter
IC2	CD4051 BCD-to-decimal decoder (SP8T rotary bilateral switch)
C1–C8	0.01 µF close tolerance capacitor
C9, C10	0.1 µF capacitor
R1	1K ¼ W 5% resistor

McGRAW-HILL

Fig. 32-18

VOLTAGE-CONTROLLED LOW-PASS FILTER

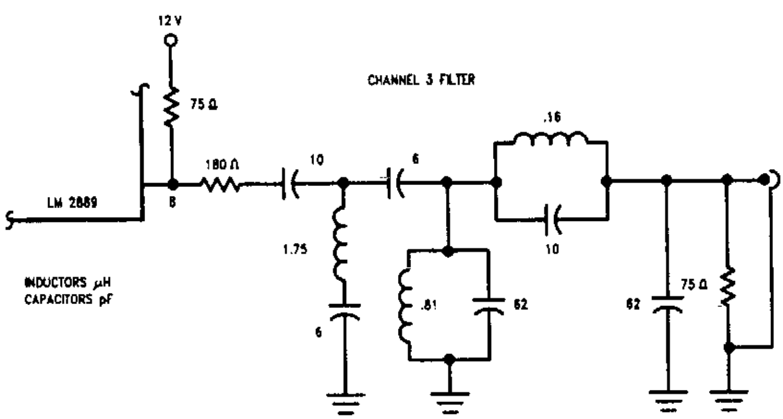


ANALOG DEVICES

Fig. 32-19

The voltage at Output A, which should be unloaded by a follower, responds as though E_s were directed to the RC filter, but the filter's break frequency were proportional to E_c [i.e,. = $E_c/(20\pi RC)$]. The frequency response has a break at f_2 and the 6-dB/octave rolloff. The voltage at Output B has the same response, up to [f_1 ($f_1 = 1/(2\pi RC)$], then levels off at a constant attenuation of $f_2/f_1 = E_c/10$. For example, if R = 8 kΩ, C = 0.002 µF, Output A has a pole at 100 Hz to 10 kHz and can be loaded. The circuit can be converted to high-pass by interchanging C and R.

VSB FILTER FOR LM2889

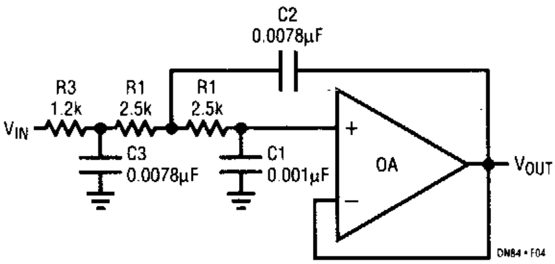


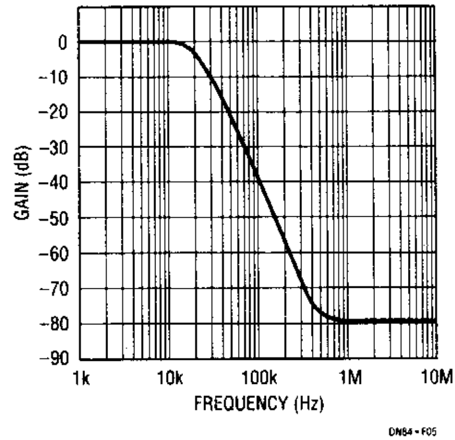
NATIONAL SEMICONDUCTOR

Fig. 32-20

This filter is for CH3, in order to get a vestigial sideband TV signal. It is designed for 75-Ω impedance levels.

20-kHz BUTTERWORTH ACTIVE FILTER



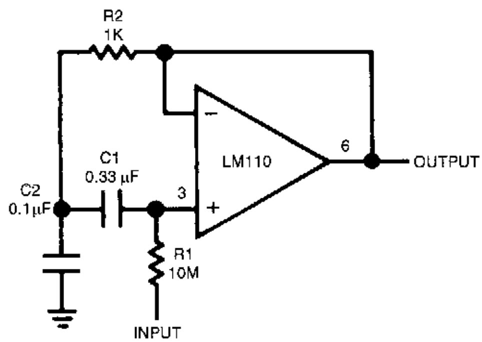


Filter Frequency Response

LINEAR TECHNOLOGY *Fig. 32-21*

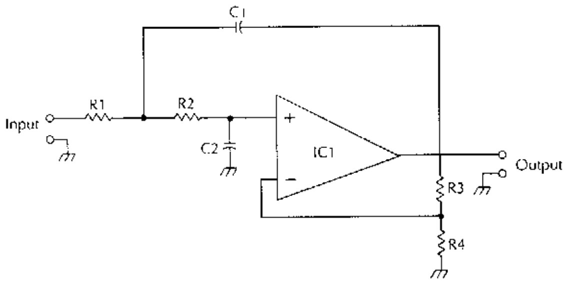
This filter will be useful for anti-aliasing or band limiting in an audio system. The op amp is a Linear Technology, LT1124, LT1355, or LT1169.

BANDPASS FILTER



NATIONAL SEMICONDUCTOR *Fig. 32-22*

SALLEN-KEY LOW-PASS FILTER



R3 and R4 set the circuit gain

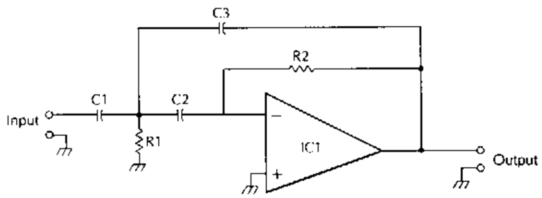
$$f_c = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

usually $C_1 = C_2$, $R_1 = R_2$, $R_3 = 0.586\,R_4$

McGRAW-HILL

Fig. 32-23

ACTIVE HIGH-PASS FILTER



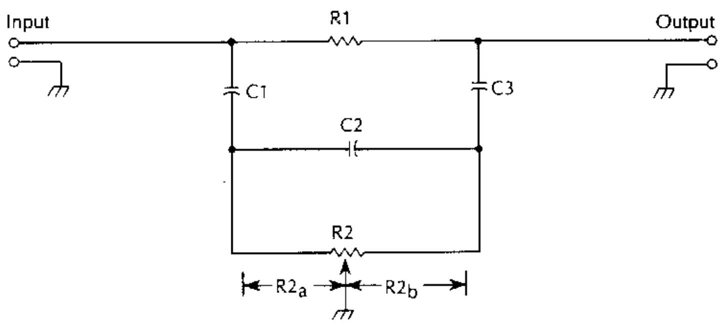
In this circuit,

$$f_{CO} = \frac{1}{2\pi RC}$$

McGRAW-HILL

Fig. 32-24

RC NOTCH FILTER



$C_1 = C_2 = C_3$
$R_1 = 6R_2$

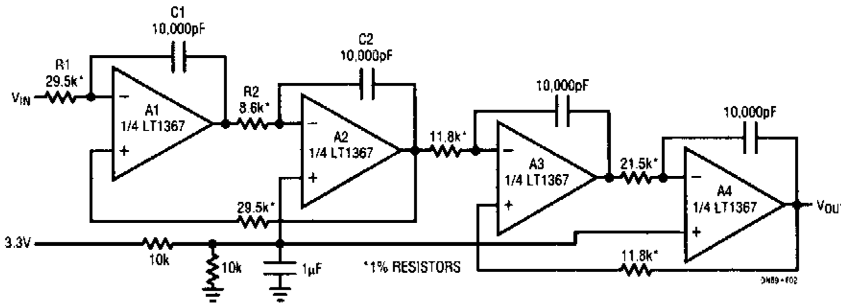
$$R_2 = R_{2a} + R_{2b}$$

$$\text{Reject frequency (notch)}, F_c = \frac{1}{2\pi C\sqrt{3R_{2a}R_{2b}}}$$

McGRAW-HILL

Fig. 32-25

1-kHz 4TH-ORDER BUTTERWORTH FILTER



LINEAR TECHNOLOGY ***Fig. 32-26***

The filter is a simplified state variable architecture consisting of two cascaded 2nd-order sections. Each section uses the 360° phase shift around the two op-amp loop to create a negative summing junction at A1's positive input. The circuit has low sensitivities for center frequency and Q, which are set with the following equations:

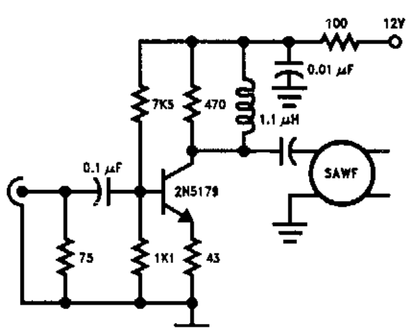
$$\omega\,0^2 = \frac{1}{(R_1 \times C_1 \times R_2 \times C_2)}$$

where,

$$R_1 = \frac{1}{(\omega 0 \times Q \times C_1)} \text{ and } R_2 = \frac{Q}{(\omega 0 \times C_2)}$$

The dc bias applied to A2 and A4, half supply, is not needed when split supplies are available. The circuit swings rail-to-rail in the passband making it an excellent anti-aliasing filter for A/Ds. The amplitude response is flat to 1 kHz then rolls off at 80 dB/decade.

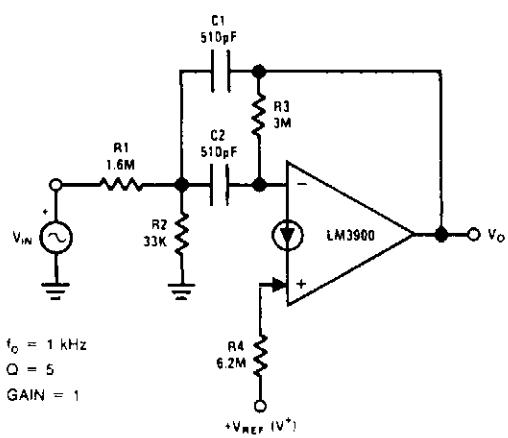
SAW-FILTER IMPEDANCE-MATCHING PREAMPLIFIER



This circuit matches a saw filter to an IF amplifier.

NATIONAL SEMICONDUCTOR ***Fig. 32-27***

ONE OP-AMP BANDPASS FILTER



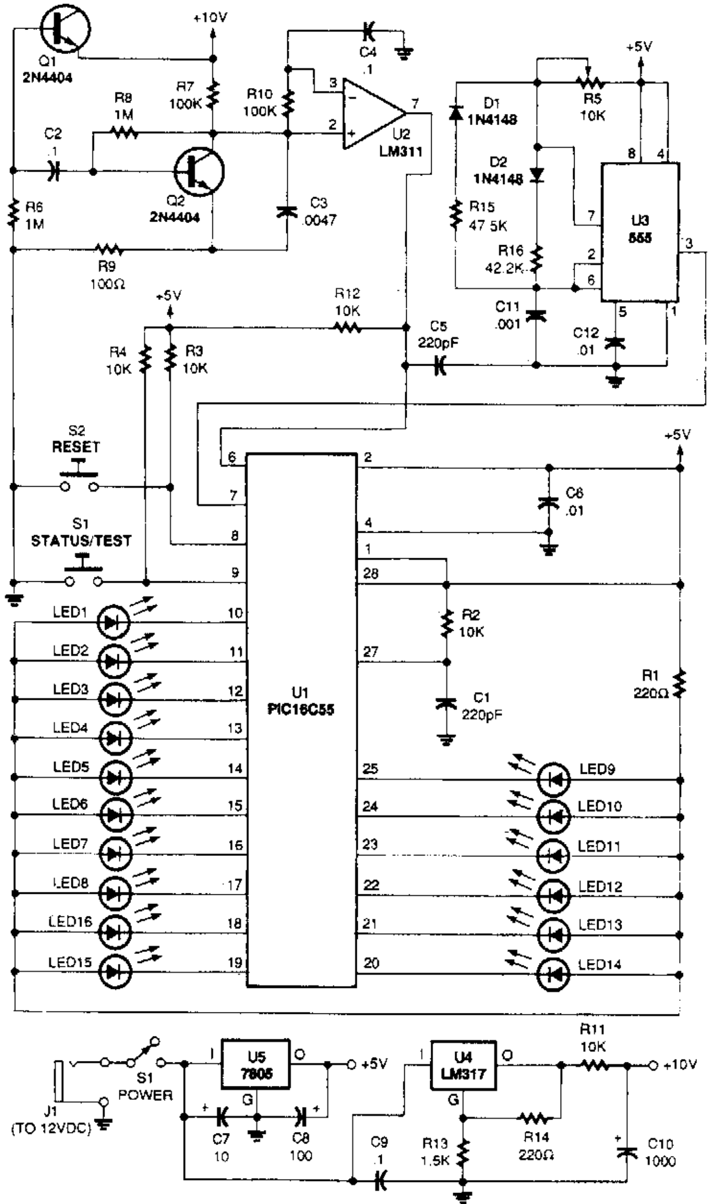
NATIONAL SEMICONDUCTOR ***Fig. 32-28***

33

Flasher Circuits and Blinker

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

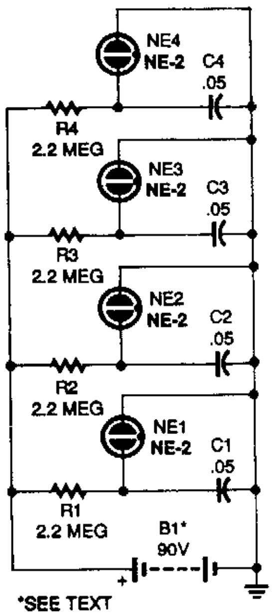
RANDOM LED STROBE



POPULAR ELECTRONICS *Fig. 33-1*

This circuit generates a random output that is translated into LED "movement" by a preprogrammed PIC16C55 microcontroller, U1. That PIC also senses and records the bias of the LED's movement. This device was originally used for an application involving psychokinesis testing where the person was asked to "think" the lights in either a clockwise or counterclockwise direction.

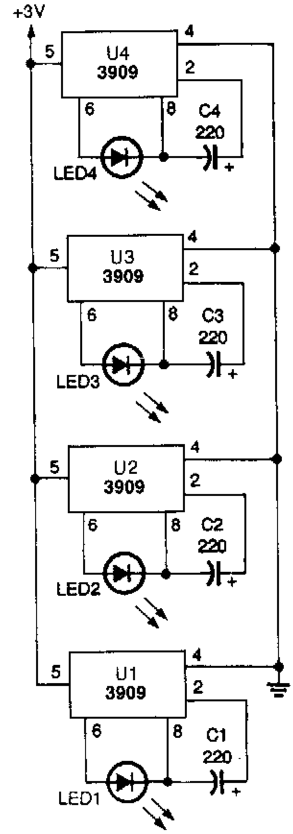
FLASHING NEON CHRISTMAS LIGHTS



POPULAR ELECTRONICS *Fig. 33-2*

This flashing set of neon Christmas lights will make an attractive decoration for any time of year. B1 is made up of ten 9-V transistor radio batteries in series. The battery life can be measured in months.

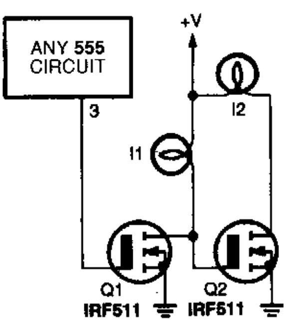
FLASHING CHRISTMAS LED DISPLAY



POPULAR ELECTRONICS *Fig. 33-3*

Using LEDs and 3909 ICs, you can make a flashing-light circuit that will run for months on two AA batteries.

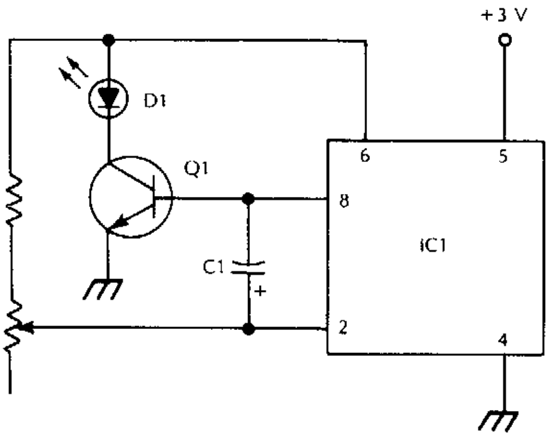
DUAL FLASHER ADD-ON FOR 555 CIRCUITS



POPULAR ELECTRONICS *Fig. 33-4*

A pair of hex FETs drive two incandescent lamps in an alternating flasher circuit. The lamps can be 12-V automotive types, etc.

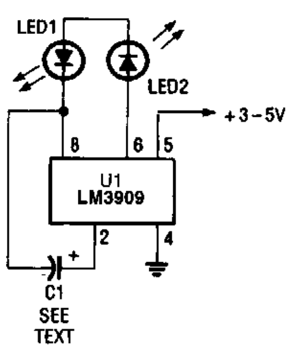
VARIABLE-FREQUENCY HIGH-POWER LED FLASHER



IC1	LM3909 LED flasher/oscillator IC
Q1	NPN transistor (2N3904, Radio Shack RS2009 or similar)
D1	LED
C1	100 µF 5 V electrolytic capacitor
R1	470 Ω ¼ W 5% resistor
R2	50 kΩ potentiometer

McGRAW-HILL *Fig. 33-5*

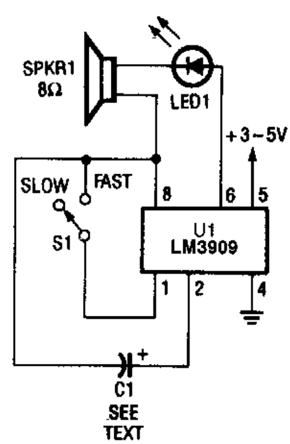
LED PULSER



POPULAR ELECTRONICS *Fig. 33-6*

In this circuit, the LM3909 is used to drive a pair of series-connected LEDs.

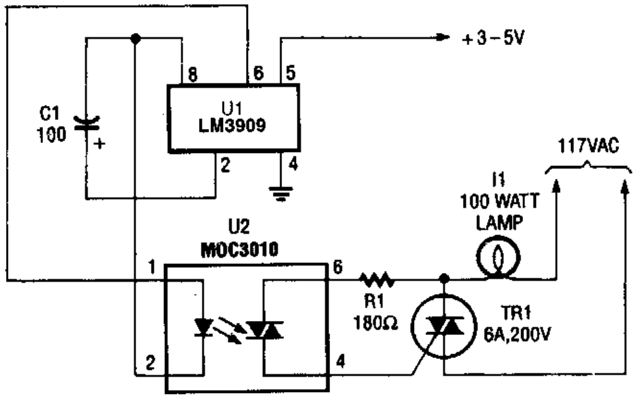
LED PULSER WITH AUDIBLE OUTPUT



POPULAR ELECTRONICS *Fig. 33-7*

The LM3090 can also be used to drive both an LED and a speaker. In this circuit, each time that LED1 blinks, SPKR1 (an 8-Ω speaker) emits a sharp click sound.

SIMPLE LAMP PULSER

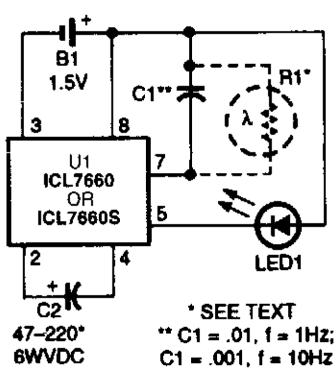


POPULAR ELECTRONICS

Fig. 33-8

Here, the LM3090 (configured as a timing oscillator) is used to control a 117-Vac lamp through an MOC3010 optoisolator/coupler.

LED FLASHER



POPULAR ELECTRONICS

Fig. 33-9

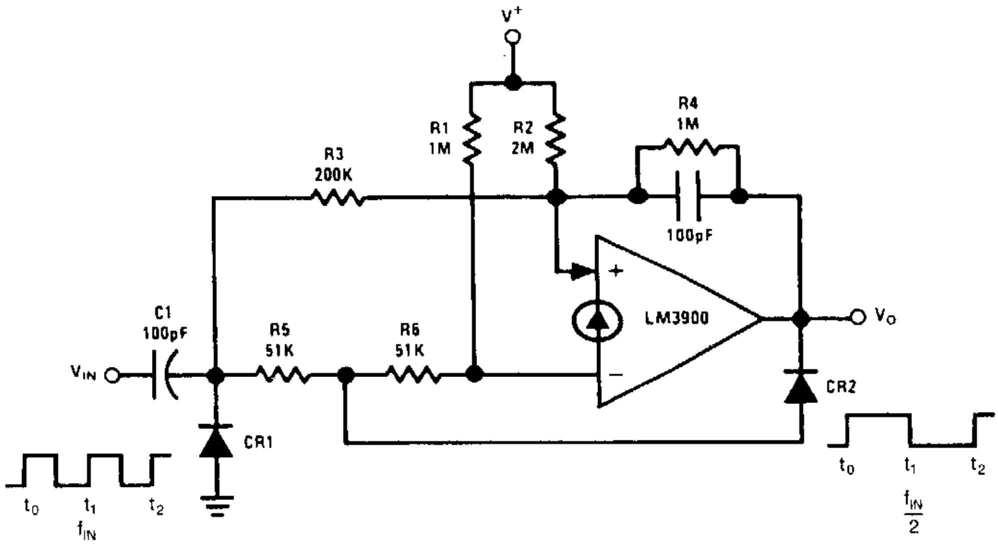
This circuit provides a low-cost way to flash an LED from a single 1.5-V source. Based on the ICL7660 dc-to-dc voltage converter, the circuit makes use of an external capacitance (C1) on the oscillator rate-control pin to decrease the charge/dump time to the desired flash rate. A dc resistance (R1) on the same pin can also be used to disable the oscillator and extend the power-cell's life. That optional dc resistance (in the form of a photoconductive cell) will shut off the oscillator in daylight.

34

Flip-Flop Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

TRIGGER FLIP FLOP

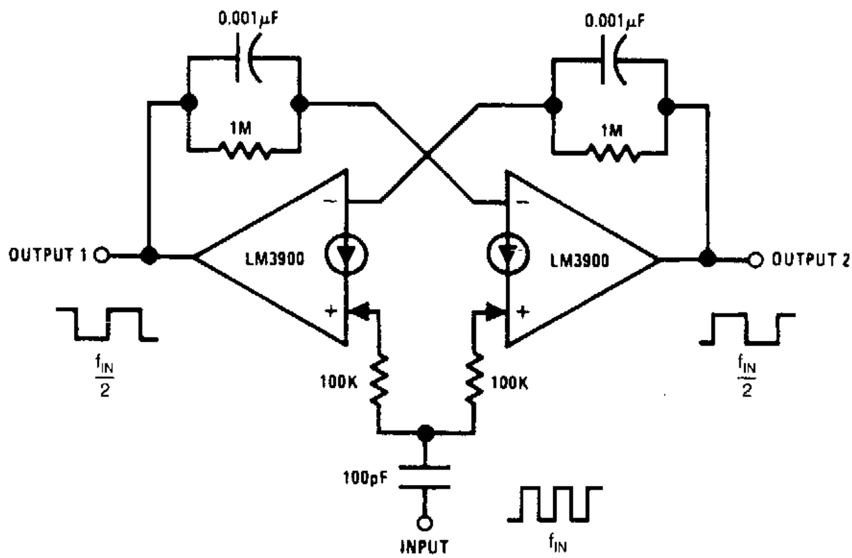


NATIONAL SEMICONDUCTOR

Fig. 34-1

Trigger flip flops are useful to divide an input frequency as each input pulse will cause the output of a trigger flip flop to change state. Due to the absence of a clocking signal input, this is for an asynchronous logic application. A circuit that uses only one amplifier is shown. Steering of the differentiated positive input trigger is provided by diode CR2. For a low-output voltage state, CR2 shunts the trigger away from the (–) input and resistor R3 couples this positive input trigger to the (+) input terminal. This causes the output to switch high. The high-voltage output state now keeps CR2 off and the smaller value of $(R_5 + R_6)$ compared with R3 causes a larger positive input trigger to be coupled to the (–) input, which causes the output to switch to the low-voltage state.

TWO-AMPLIFIER FLIP FLOP



NATIONAL SEMICONDUCTOR

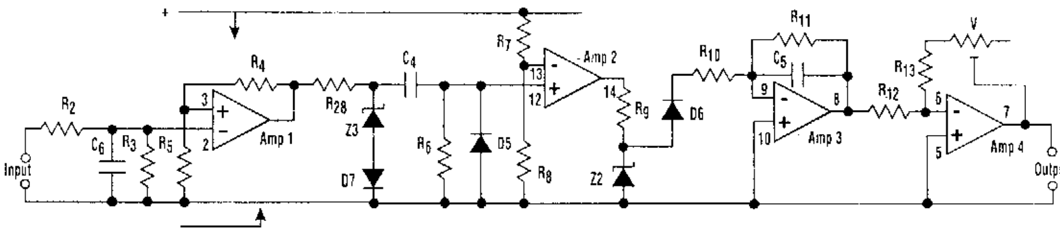
Fig. 34-2

35

Frequency-to-Voltage Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

INDUSTRIAL FREQUENCY-TO-VOLTAGE CONVERTER



ELECTRONIC DESIGN *Fig. 35-1*

Control and process equipment often require the indication of frequency (speed or rate) of linear or rotary mechanical movement. Motion can be detected using various pulse-generating pickups and proximity detectors that output ac or dc pulsed signals.

This industrial converter can serve in a wide variety of applications. The circuit operates around a quad-FET input op amp and is designed to be self-contained or run from a bipolar supply. The input signal of dc pulses or ac waveforms is applied to R2.

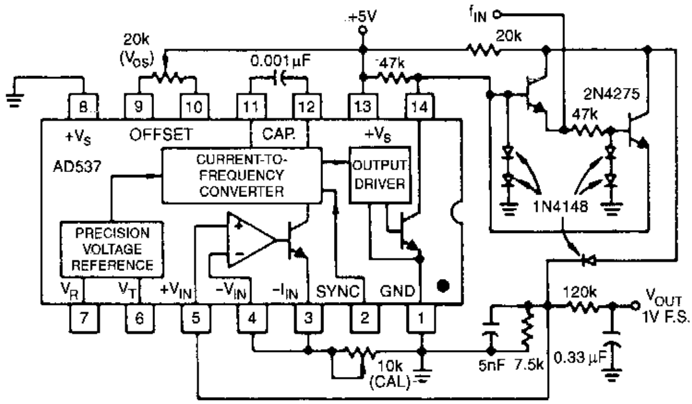
Amplifier 1, which acts as a Schmitt trigger, supplies a squarewave output of fixed amplitude to C4. Resistor R3 and capacitor C6 filter out input noise, and R4 and R5 determine the switching levels, and Zener diode Z3 sets the amplitude.

Amplifier 2 gives a fixed-duration pulse on the positive transition of C4, with a time constant set by C4 and R6 and the switching level set by R7 and R8. Resistor R9 and Zener diode Z2 fix the amplitude of the pulses and amplifier 3 integrates them via R10 and C5. Diode D6 blocks negative integration and R11 discharges C5 with a long-time constant.

Hence, the dc output of amplifier 3 is proportional to the frequency applied to the input. Amplifier 4 inverts and buffers the negative output of amplifier 3 and provides amplitude adjustment voltage.

The complete circuit is linear and sufficiently accurate providing that C4 is chosen to give a pulse duration less than the maximum input frequency and that R11 > R10.

FREQUENCY-TO-VOLTAGE CONVERTER



ANALOG DEVICES ***Fig. 35-2***

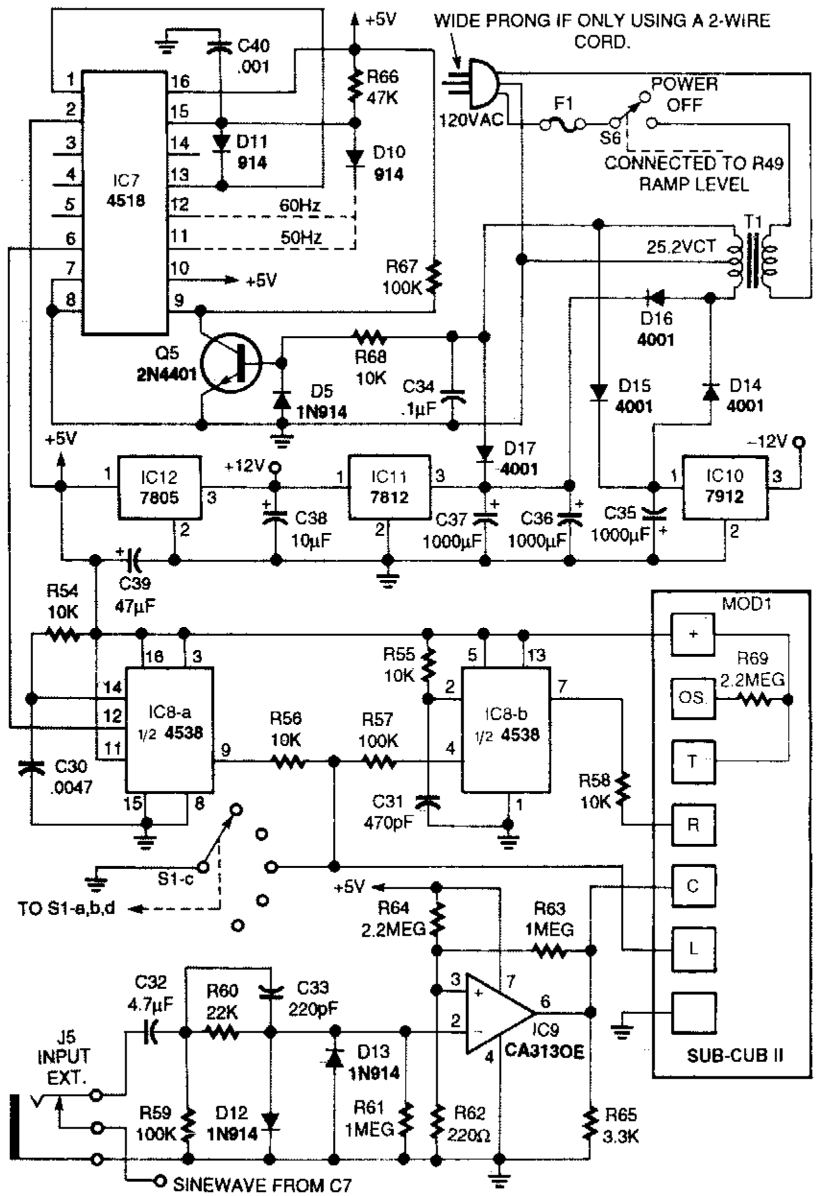
The AD537 can also be used to perform frequency-to-voltage conversion. The transistor pair shown here operates as an exclusive-or gate to perform the phase comparison. It locks onto the input frequency within two cycles. The configuration requires only 3 mA for frequencies up to 10 kHz. In most situations, an output buffer will be required to unload the filter. Use 0- to 5-V pulses or square waves with 40-µs minimum pulsewidth.

36

Function Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

FUNCTION GENERATOR

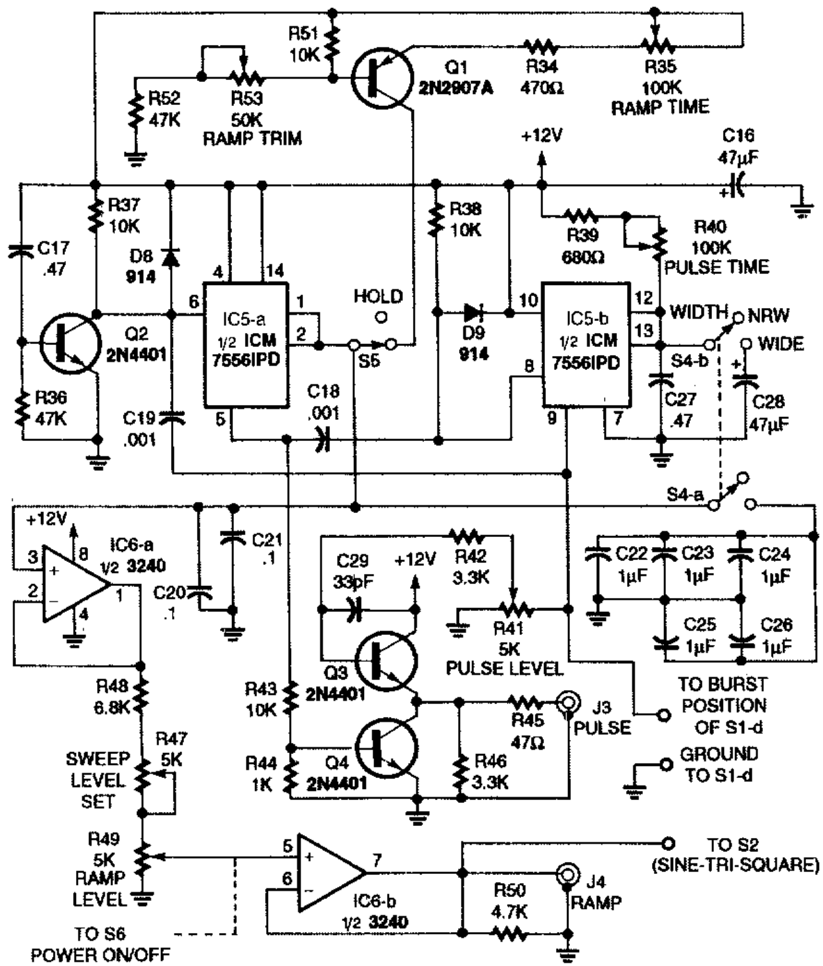


THE COUNTER MODULE (MOD1) has a 0.35-inch high, six-digit liquid crystal display. Pulses at 1-second intervals are derived from the AC power line which has a typical accuracy of 99.99 %.

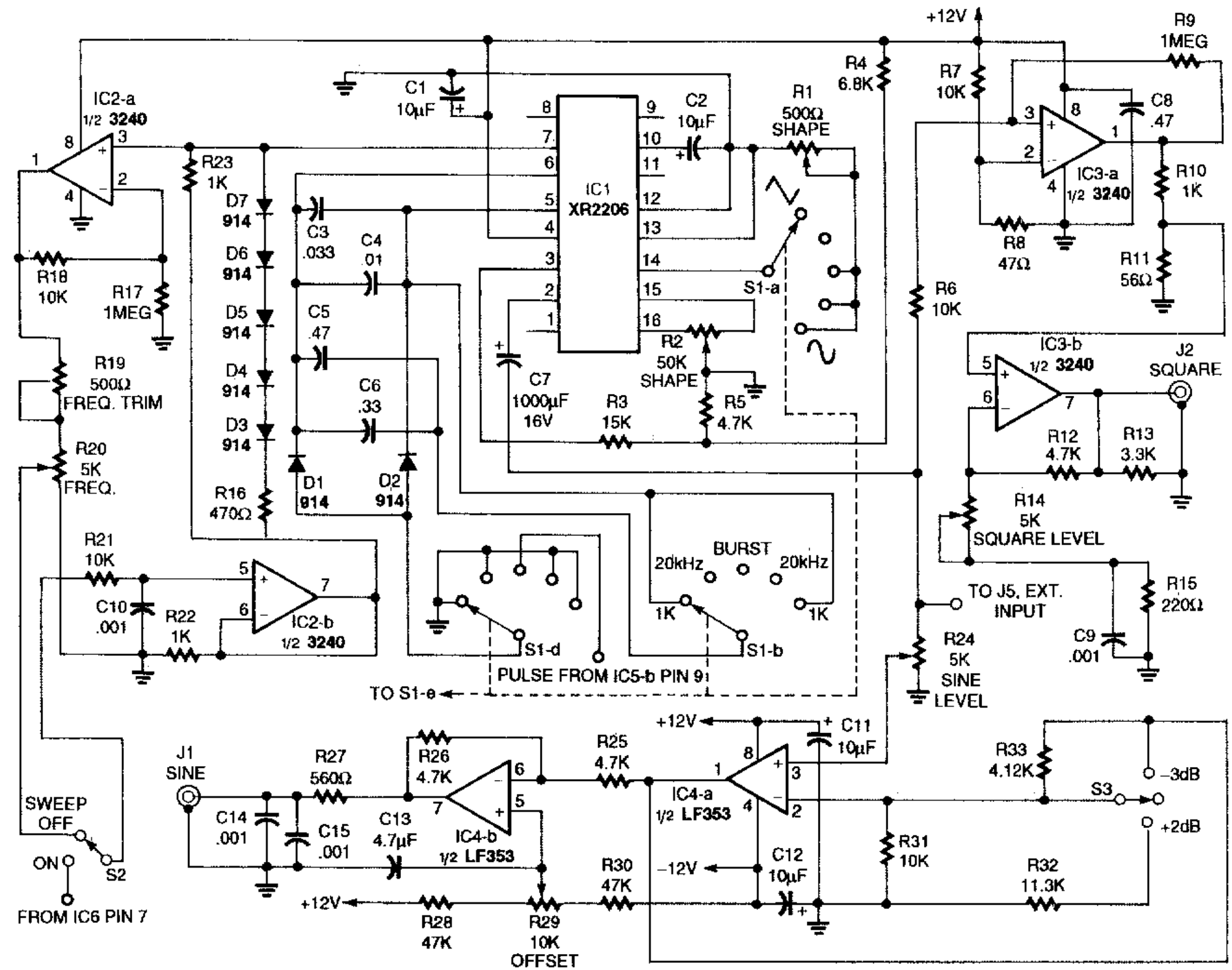
ELECTRONICS NOW *Fig. 36-1*

These three circuits make up an audio frequency function generator and can be individually used for custom applications.

FUNCTION GENERATOR (*Cont.*)

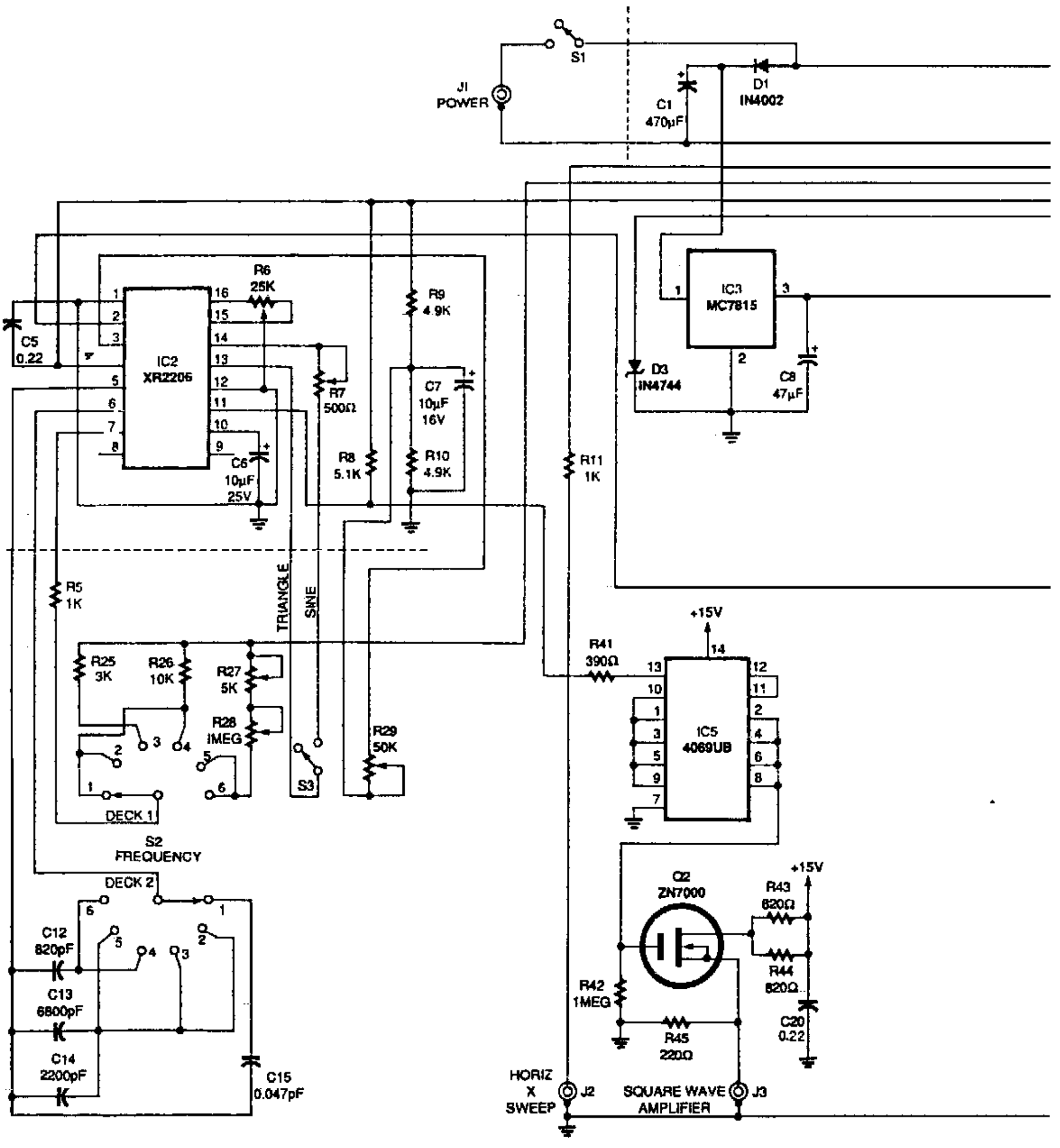


A DUAL CMOS TIMER'S (IC5) output triggers another timer's input. Once Q2 starts IC5-a on initial power-up, the circuit continues to oscillate.



AN XR 2206 FUNCTION-GENERATOR CHIP provides a triangle output at pin 2 when S1-a is open.

SWEEP/FUNCTION GENERATOR



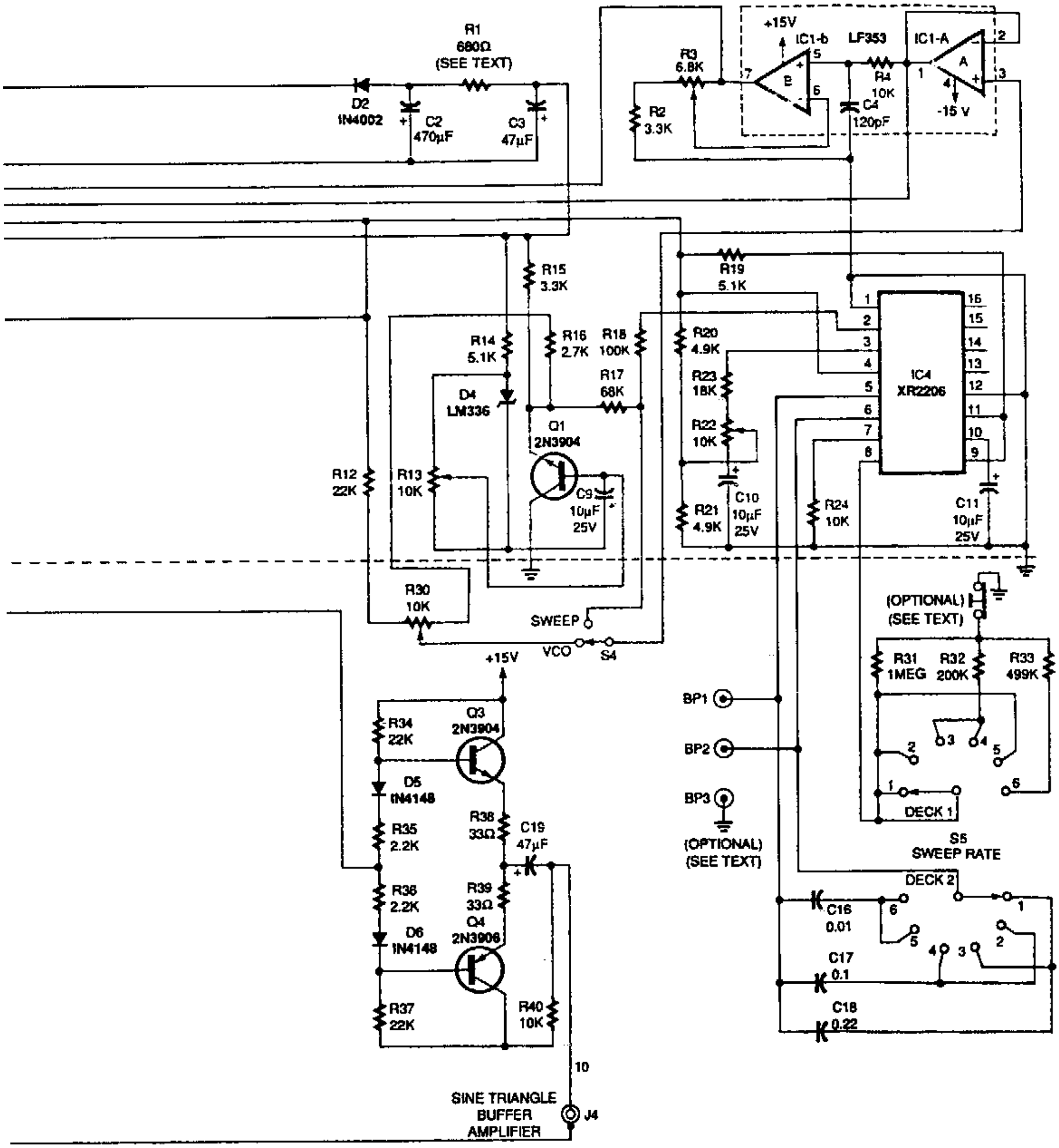

R1
680Ω
(SEE TEXT)
D2
1N4002
C2
470µF
C3
47µF
+15V
IC1-b
LF353
IC1-A
R3
6.8K
R4
10K
C4
120pF
R2
3.3K
-15 V
R15
3.3K
R19
5.1K
R14
5.1K
R16
2.7K
R18
100K
R17
68K
R20
4.9K
R23
18K
R22
10K
D4
LM336
Q1
2N3904
IC4
XR2206
R12
22K
R13
10K
C9
10µF
25V
C10
10µF
25V
R21
4.9K
R24
10K
C11
10µF
25V
R30
10K
SWEEP
VCO
S4
+15V
(OPTIONAL)
(SEE TEXT)
R31
1MEG
R32
200K
R33
499K
BP1
BP2
BP3
(OPTIONAL)
(SEE TEXT)
Q3
2N3904
R34
22K
D5
1N4148
R35
2.2K
R38
33Ω
C19
47µF
R36
2.2K
R39
33Ω
Q4
2N3906
D6
1N4148
R37
22K
R40
10K
DECK 1
S5
SWEEP RATE
DECK 2
C16
0.01
C17
0.1
C18
0.22
SINE TRIANGLE
BUFFER
AMPLIFIER
J4

Fig. 36-2

SWEEP/FUNCTION GENERATOR (*Cont.*)

TABLE 1
FUNCTION GENERATOR CHARACTERISTICS

Waveform output	Maximum P-to-P	Frequency	Conditions
Sine (1)	5V	10 Hz-100 kHz	1 V@800 kHz
Triangle (1)	8 V	10 Hz- 50 kHz	1 V>500 kHz
Square (2)	5 V		Positive output DC-coupled, ground ref: rise/fall >50 ns
Ramp (3)			Descending, 6 rates

(1) Output level variable frim min. to max.
(2) Output level not adjustable.
(3) X and Y amplitude internally adjustable.

TABLE 2
SWEEP RANGES OF THE FUNCTION GENERATOR

Switch	Condition	Frequency range
1	Preset	20Hz to >2kHz
2	Preset	<400Hz to >10kHz
3	Preset	<1kHz to >25kHz
4	Preset	5kHz to >100kHz
5*	Resistance tuned	2kHz to 100kHz
	Resistance & VCO tuned	<10Hz to >100kHz
6*	Resistance tuned	<40kHz to >800kHz
	Resistance & VCO tuned	<100Hz to >800kHz

* Ranges show for positions 5 and 6 represent the total tuning range of the function generator and do not imply one continuous sweep.

Both IC2 and IC4 are Exar XR2206 monolithic function generators; IC4 functions as a ramp generator, and IC2 functions as a generator of sine, triangular, and square waveforms. Dual operational amplifier IC1 produces a scaled, level-shifted ramp output that is capable of deflecting an oscilloscope's horizontal sweep.

Any frequency of interest along the horizontal axis of an oscilloscope that is coupled to this function generator can be measured with an external frequency counter by manually tuning the function generator's VCO instead of sweeping it. The performance characteristics of the sweep/function generator are summarized in the Table.

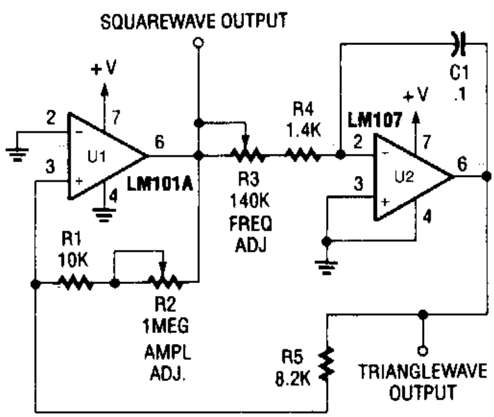
The generator's sweep rate and frequency can be set by front-panel rotary six-position switches, Sweep Rate Switch S5 and Frequency Switch S2. The VCO control R30 manually tunes the VCO. Table 2 lists the sweep ranges of the function generator. Sweep ranges not covered in ranges 1 to 4 can be set up as required on positions 5 and 6. Selecting the VCO setting on the front panel toggle switch S4 permits tuning any fixed frequency within the total frequency range of the instrument with both frequency switch S2 and VCO control R30.

The sweep rate or duration of the sweep ramp is selected by the rotary six-position Sweep Rate Switch S5. Table 3 lists the sweep rate durations for each of the six positions. Longer periods should be used for lower frequency sweeps.

TABLE 3
SWEEP RATE OR DURATION

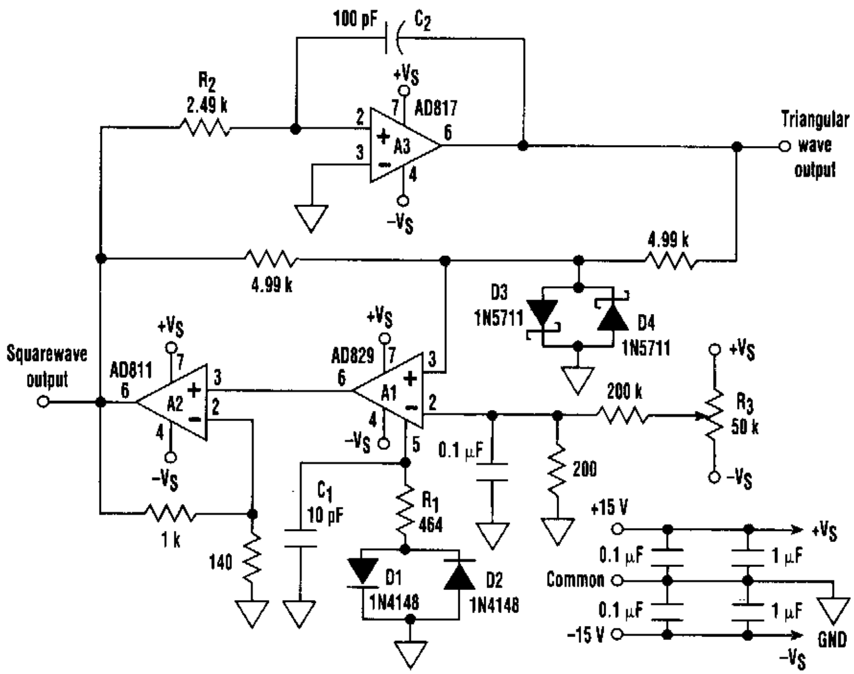
Sweep position	Period (milliseconds)
1	~130
2	~ 60
3	~ 30
4	~ 15
5	~ 6
6	~ 3

SIMPLE FUNCTION GENERATOR



POPULAR ELECTRONICS *Fig. 36-3*

ACCURATE, STABLE FUNCTION GENERATOR



ELECTRONIC DESIGN

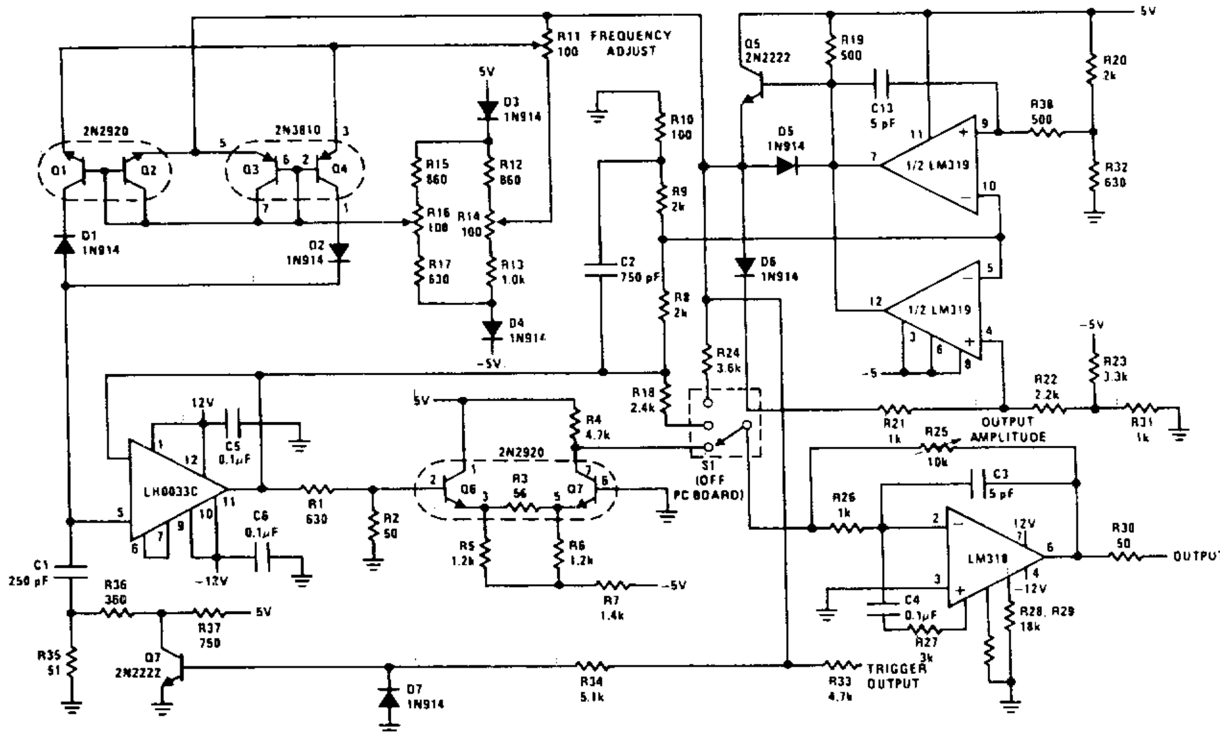
Fig. 36-4

Supply-limited oscillators usually are sensitive to temperature and power-supply changes, are never symmetrical, and don't operate at high frequencies because the amplifier's output is saturated when it reaches the supply lines.

The circuit shown, a function generator, can alleviate these problems. Its square-wave output boasts a rapid rise time, quick settling time, and an amplitude that's temperature insensitive. Also, its triangular output waveform features a perfectly constant rate of change throughout its range.

Amplifier A1 together with A2 generates a stable +10 V. This signal, which is integrated using A3, C2, and R2, makes a negative-going ramp. When the peak output of A3 equals −10 V, the output of A1 and A2 change state and the A3's output ramps up. When A3's output equals +10 V, the outputs of A1 and A2 change state again and new cycle starts.

WIDE-RANGE FUNCTION GENERATOR



NATIONAL SEMICONDUCTOR *Fig. 36-5*

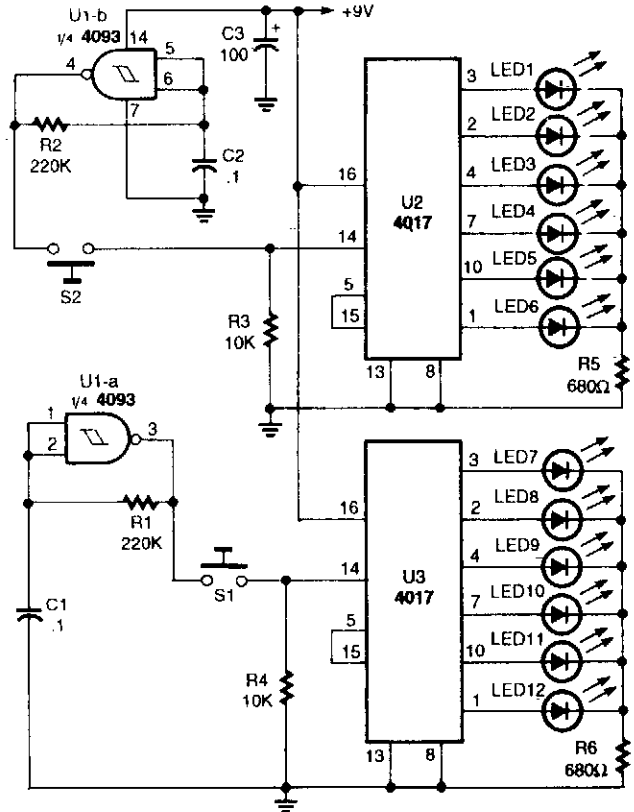
The sine, square, triangle function generator is exceptionally useful. Various IC circuits have been published for generating square and triangle waveforms in an attempt to duplicate the general-purpose function generator. However, these simple circuits are usually limited to about 10 kHz and have no sine-wave output. The function generator shown here provides all three waveforms and operates from below 10 Hz to 1 MHz with usable output to about 2 MHz.

37

Game Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC CRAPS GAME

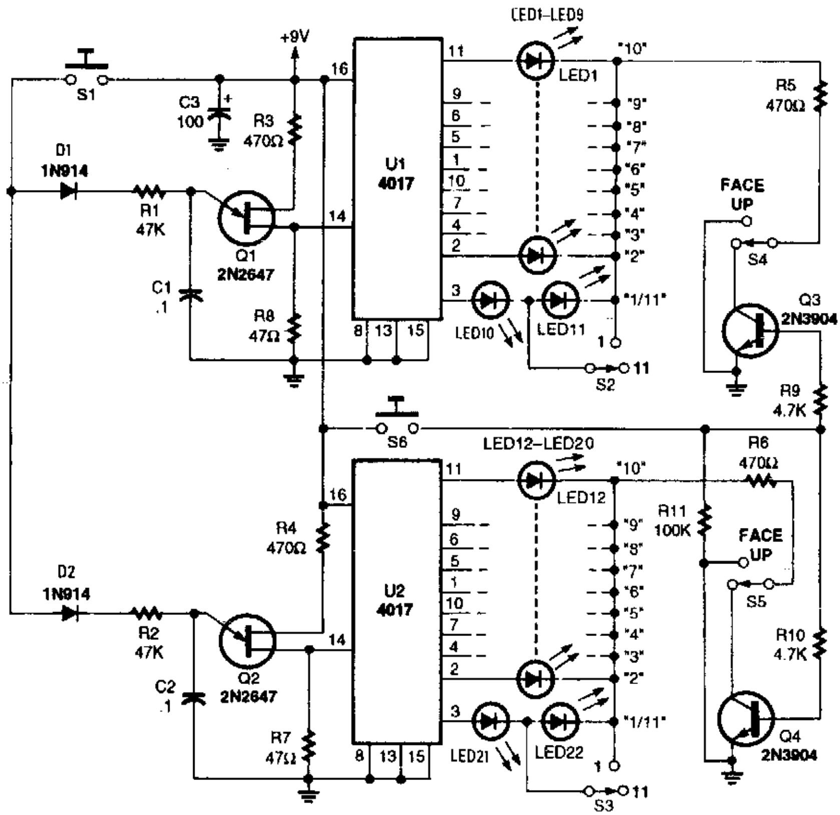


POPULAR ELECTRONICS *Fig. 37-1*

Two gates of a 4093 quad, 2-input NAND, Schmitt-trigger CMOS IC are connected in astable-oscillator circuits as clocks. The two 4017 ICs have six LEDs connected to its first six outputs. As the clock pulses enter pin 14 of the 4017s, the ICs count from one to six over and over as long as the clock pulses are present. When S1 and/or S2 are released, one of the LEDs in each circuit will remain on, indicating a number from one to six.

The circuit is set up so that you can roll the dice together by pressing S1 and S2 at the same time, or roll each die one at a time.

21 GAME



POPULAR ELECTRONICS *Fig. 37-2*

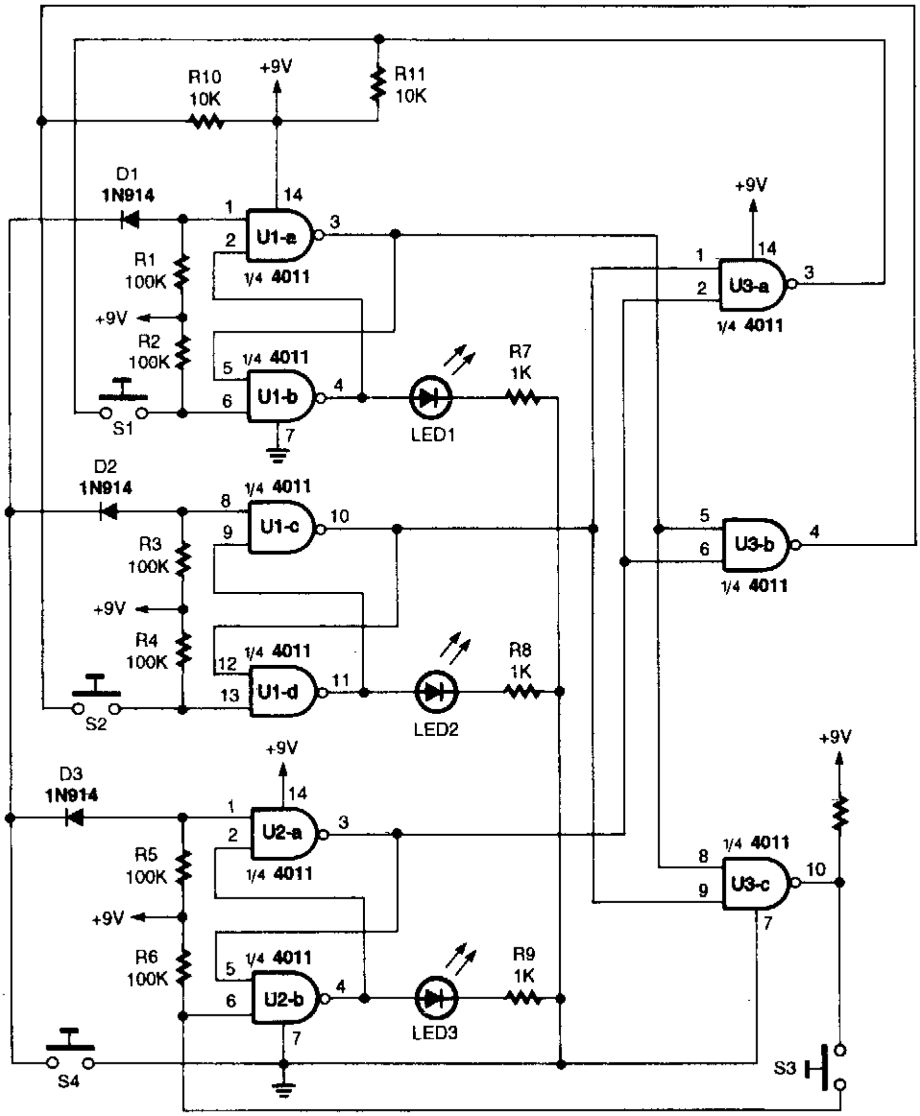
Two 2N2647 unijunction transistors serve as the clock generators for the two 4017 ICs. A single "deal" push-button switch, S1, operates both clock generators at the same time. Diodes D1 and D2 isolate the two clock circuits, allowing S1 to operate both.

The 4017 counter/readout circuits are identical in circuitry and operation. As long as clock pulses enter pin 14 of each 4017, the ICs count from 1 to 10 over and over until the clock pulses stop. When S1 is released, the clock pulses stop and one LED from each IC remains on to indicate a card with a number value of 1 (1 or 11) to 10.

The position of switches S2 and S3 determines whether the number 1 ("Ace") output of the 4017s count as an 11 or a 1. Both S2 and S3 can be switched in either position before or after the cards are played.

The cards can be played either face up or face down. When switches S4 and S5 are in the position shown in the figure, the cards are dealt face down. Transistors Q3 and Q4 are turned off in this position and no current can flow through the LEDs. Pressing S6 turns both transistors on, lighting the LEDs.

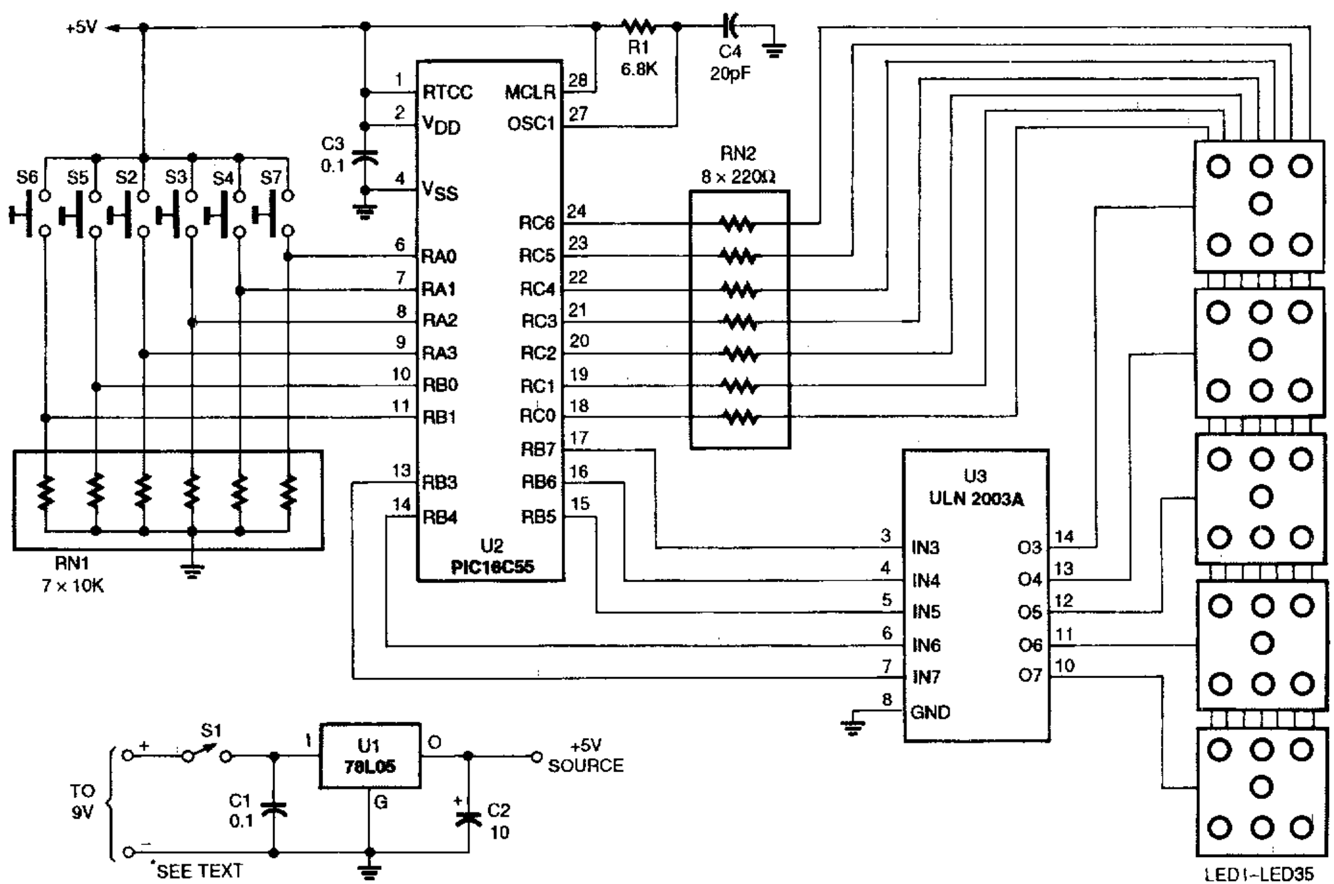
FIRST-RESPONSE MONITOR I

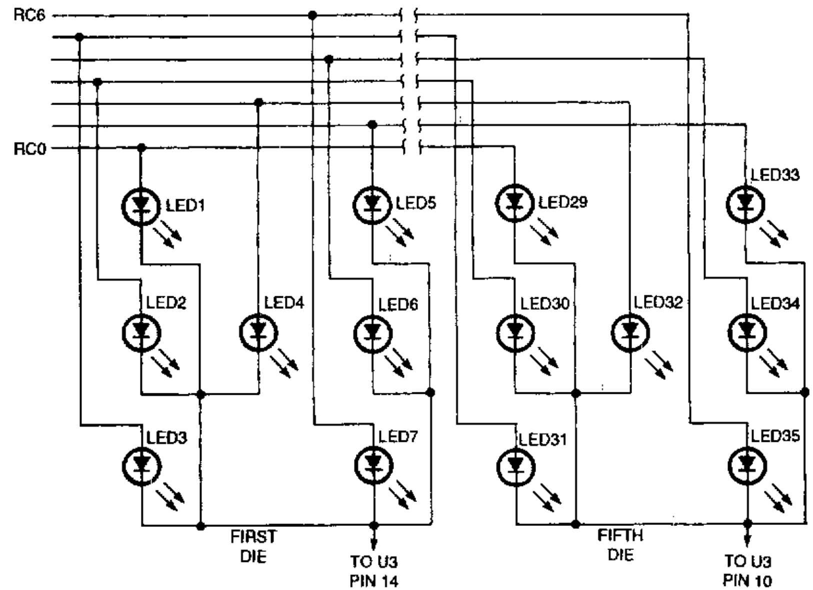


POPULAR ELECTRONICS *Fig. 37-3*

Three interlocked flip-flops enable the detection of the first input. S1, S2, and S3 are inputs. Analog switches controlled by logic gates, or other logic circuitry could be sub-statement for S1, S2, and S3.

Z-DICE GAME


+5V
R1
6.8K
C4
20pF
RTCC
MCLR
VDD
OSC1
C3
0.1
VSS
S6
S5
S2
S3
S4
S7
RA0
RA1
RA2
RA3
RB0
RB1
RB3
RB4
RC6
RC5
RC4
RC3
RC2
RC1
RC0
RB7
RB6
RB5
U2
PIC16C55
RN1
7 × 10K
RN2
8 × 220Ω
U3
ULN 2003A
IN3
IN4
IN5
IN6
IN7
GND
O3
O4
O5
O6
O7
LED1–LED35
TO
9V
S1
*SEE TEXT
C1
0.1
U1
78L05
I
O
G
C2
10
+5V
SOURCE



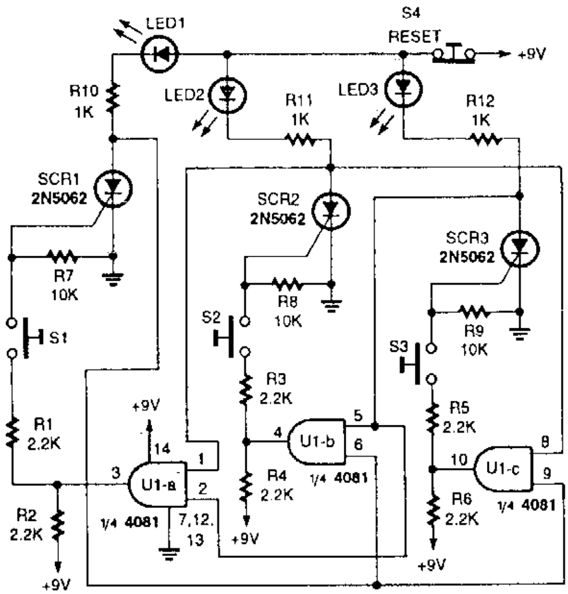
POPULAR ELECTRONICS

Fig. 37-4

Using a microcontroller (U2) keeps the parts count and the cost of this 5-dice LED display relatively low. Z-dice uses five clusters of seven LEDs to represent the marks or "pips" on five dice. Buttons below each of the LED dice let the player mark a die to be rolled on the next throw. Marked dice show up as dimmed LEDs. Pressing the button to the right of the display rolls the marked dice. If the player changes his or her mind about rolling a particular die before pressing the roll button, he or she can unmark it by pressing its button a second time. If no dice are marked at the time the player presses the roll button, then all of the dice are marked to be rolled. A second press starts them rolling, animating the LEDs of the marked dice for a second or so before displaying the results of the roll. Z-Dice doesn't count rolls or keep score, so it's still up to the players to make sure that nobody cheats!

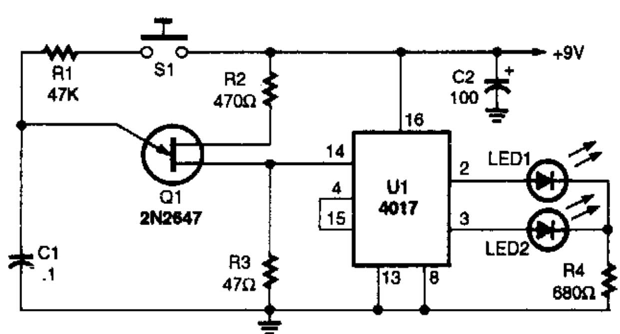
This diagram shows the wiring details of the dice display. For space and simplicity, only the first and last dice are shown. A programmed microcontroller is needed for this circuit. Refer to the original article for software.

THREE-INPUT FIRST-RESPONSE MONITOR



POPULAR ELECTRONICS *Fig. 37-5*

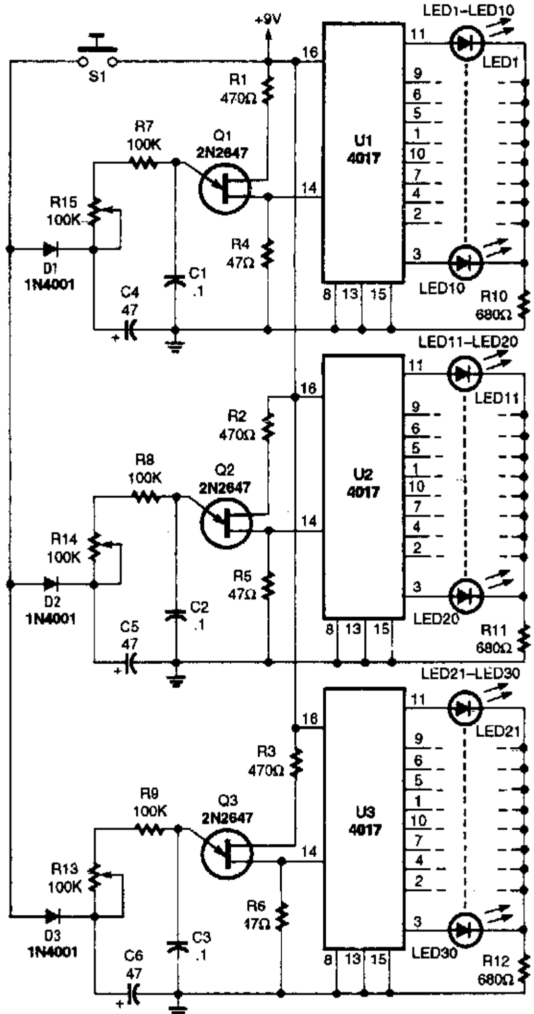
ELECTRONIC COIN TOSS



POPULAR ELECTRONICS *Fig. 37-6*

Integrated circuit U1 is connected in a two-stage counter circuit that counts "one-two" over and over as long as clock pulses enter pin 14 of the 4017. When the clock pulses stop, one of the LEDs will remain on, indicating the last even or odd count. Designate one LED as "heads" and the other as "tails" and you have an electronic coin flipper.

ELECTRONIC ONE-ARM BANDIT



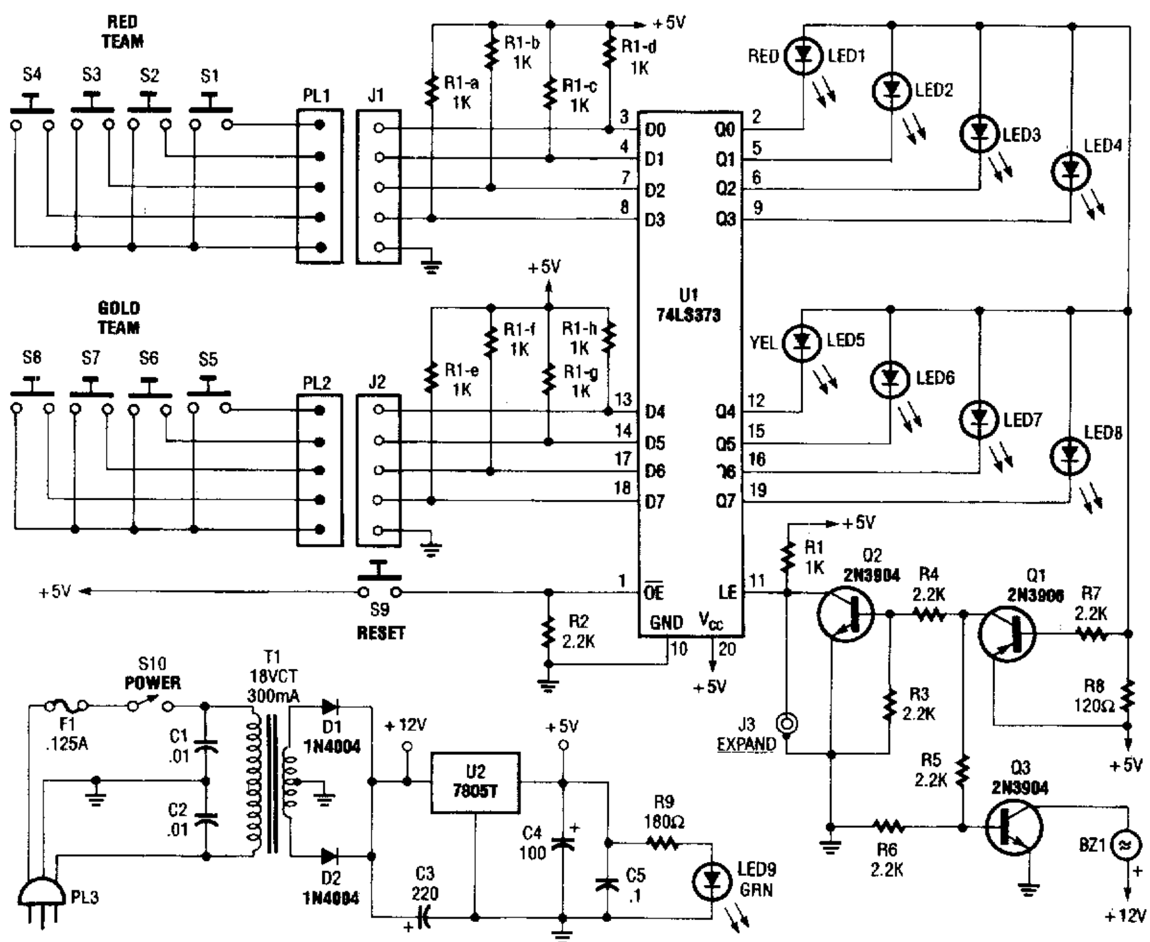
POPULAR ELECTRONICS *Fig. 37-7*

The one-arm bandit circuit is made up of three clock circuits and three counter/readout circuits. A single roll switch, S1, turns on all three clocks at the same time. When S1 is closed, capacitors C4, C5, and C6 are charged through D31, D32, and D33 to about 8 V. After S1 is released, the three clocks run, taking energy from the three charged capacitors. As the capacitors discharge, the three clocks begin to slow down, producing the effect of the drums in a mechanical bandit slowing to a stop.

The 4017's 10-output LEDs can be numbered or designated as apples, cherries, bells, wild cards, or anything you like to make the game more interesting. Additional logic circuitry can be added to the 4017 outputs to sound an alert or turn on a light when any three numbers or output items match.

Three potentiometers, R12, R13, and R14, can be varied for each roll to change the clock's frequency and the roll rate.

DIGITAL "FIRST-TO-RESPOND" BOX



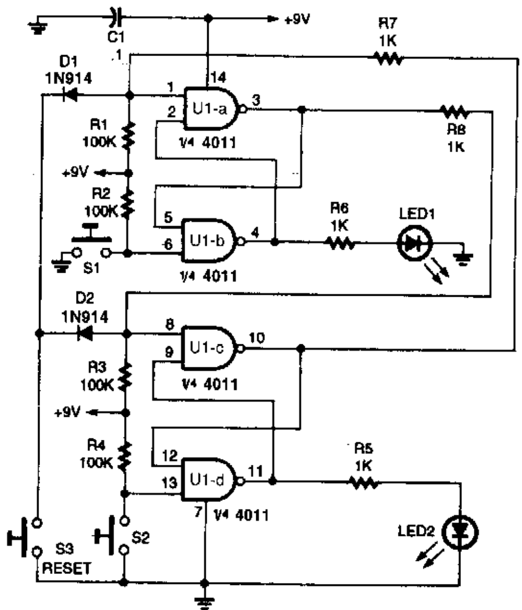
POPULAR ELECTRONICS

Fig. 37-8

This device is useful for quizzes and games to determine first response. U1 is an octal D type latch IC, an 74LS373.

When a button is pushed, this circuit lights the corresponding LED. Q1 conducts, sounding an alarm (BZ1) connected to driver Q3, and Q1 supplies bias to Q2, disabling the rest of the latches in U1.

FIRST-RESPONSE MONITOR II

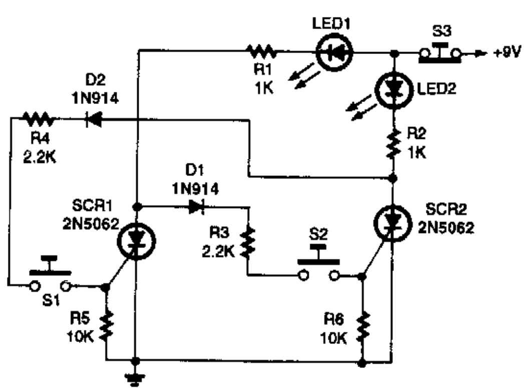


POPULAR ELECTRONICS

Fig. 37-9

Two interlocking flip flops are used to detect the first of two inputs. S1 and S2 are input devices, but a logical-level signal can be substituted.

ANALOG FIRST-RESPONSE MONITOR

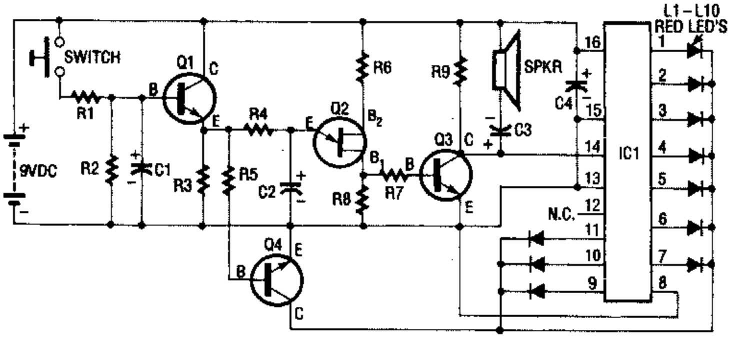


POPULAR ELECTRONICS

Fig. 37-10

The analog first-response monitor is built around a pair of cross-coupled SCRs, each of which receives its gate trigger current from the anode of the other SCR.

WHEEL OF FORTUNE



C1,C3	10 µF Capacitor
C2	1 µF Capacitor
C4	47 µF Capacitor
IC1	MC14017BCP
L1-L10	Jumbo Red LEDs
Q1,Q3,Q4	2N3904 Transistor
Q2	MU10UJT Transistor
R1,R5	33K Resistor
R2	2.2 Meg Resistor
R3	82K Resistor
R4	47K Resistor
R6	2.2K Resistor
R7	390 ohm Resistor
R8	100 ohm Resistor
R9	680 ohm Resistor
S1	Pushbutton Switch

POPULAR ELECTRONICS ***Fig. 37-11***

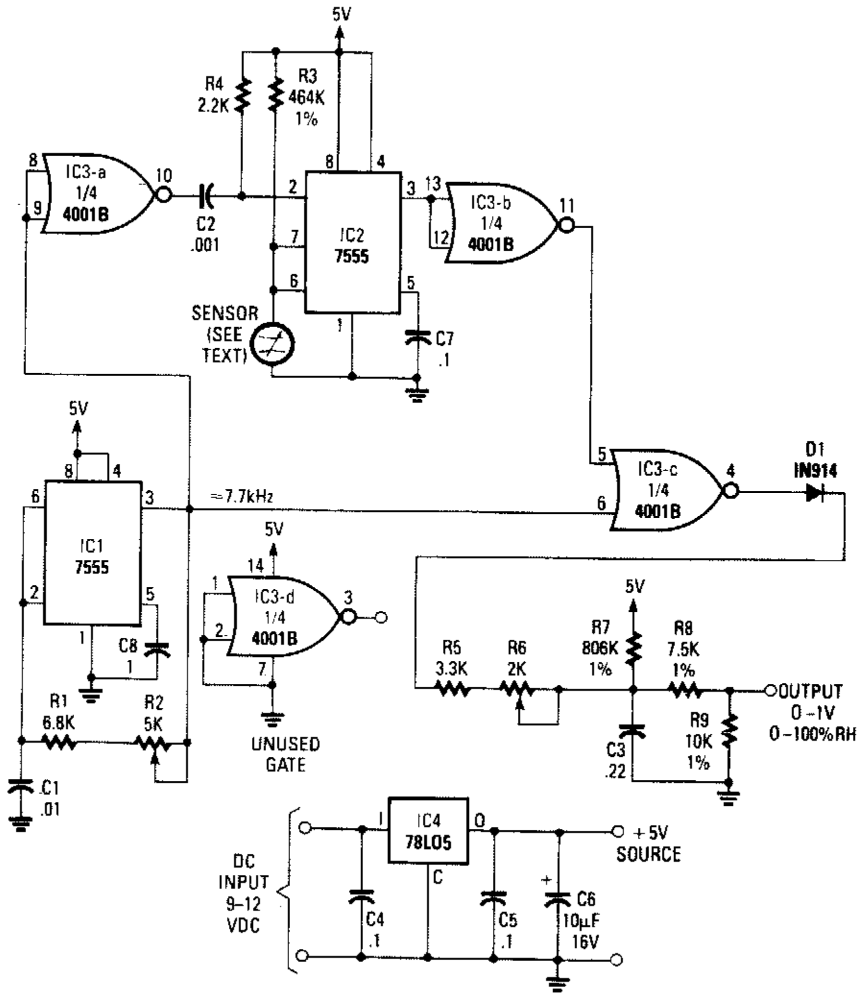
The oscillation of Q2 is amplified by Q3 and fed to Johnson counter IC1. The output of IC1 drives the LEDs in sequence to give the impression of a spinning red ball.

38

Humidity Sensor Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

HUMIDITY MONITOR

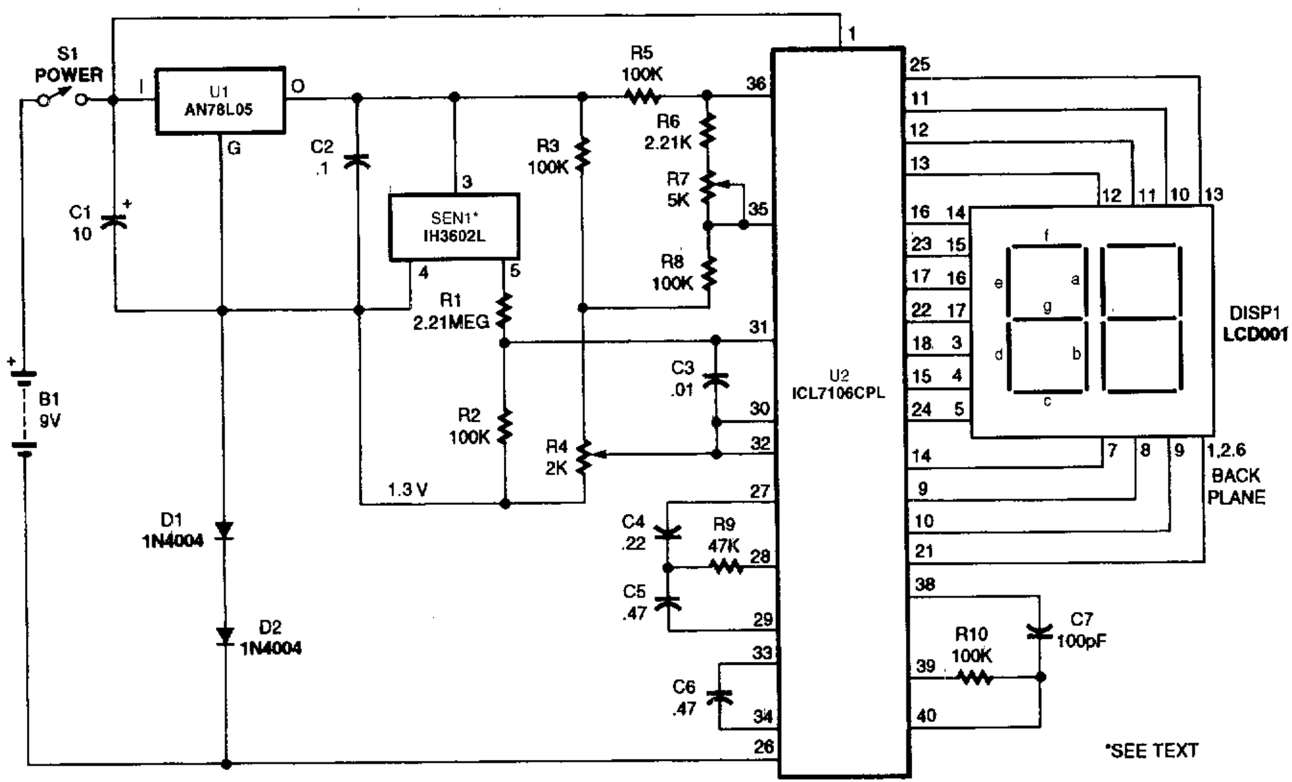


RADIO-ELECTRONICS

Fig. 38-1

This circuit uses a Phillips capacitive humidity sensor that has a ΔC variation of 45 pF over 0 to 100 pF RH. IC2 is an oscillator whose frequency is determined by the RH sensor. It is compared to fixed oscillator, and the difference frequency is taken by IC3C and rectified, outputting a 0- to 1-V signal for RH between 0 and 100%.

DIGITAL RELATIVE HUMIDITY GAUGE



POPULAR ELECTRONICS

Fig. 38-2

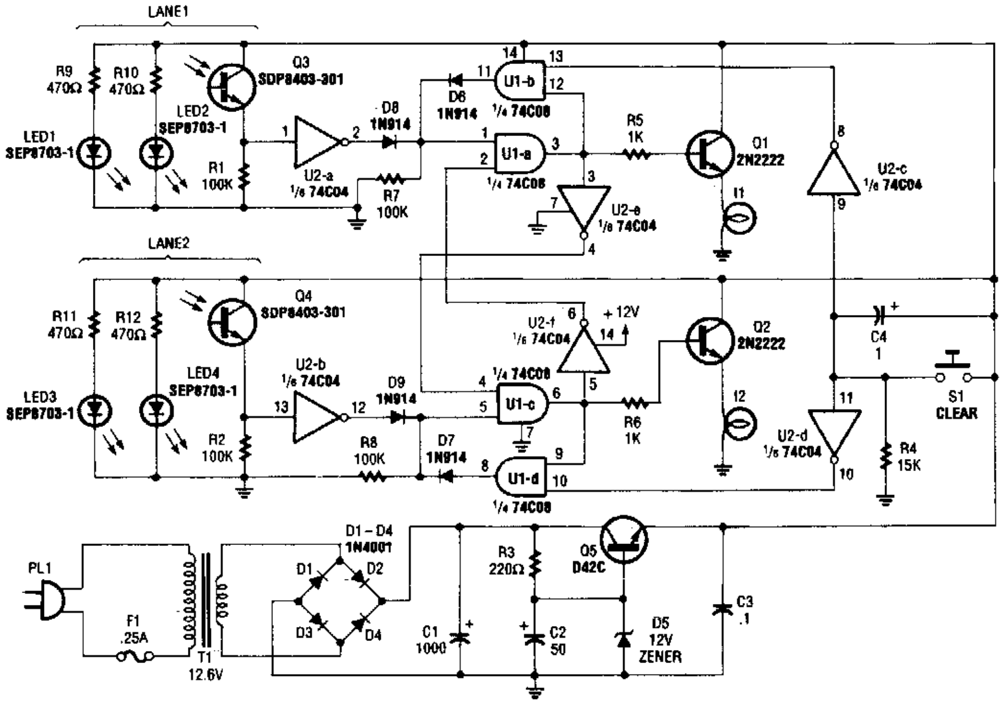
Sensor SEN1 outputs a dc voltage that varies linearly with relative humidity. This dc voltage is fed through R1 and R2 to A/D converter chip U2. Zero set is performed with R4. The LCD display is calibrated with R7 to read 0 to 100 percent.

39

Indicator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

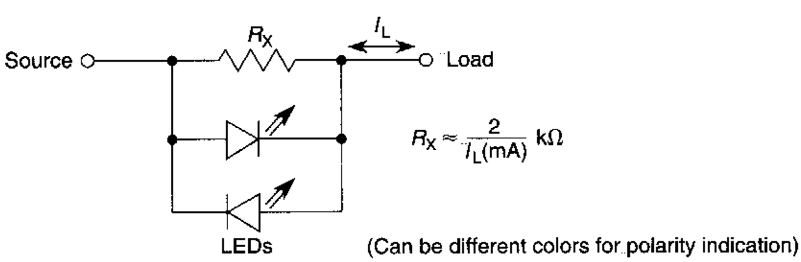
MODEL CAR DERBY WINNER INDICATOR



POPULAR ELECTRONICS *Fig. 39-1*

This derby-winner indicator uses infrared emitters and sensors to detect a car crossing the finish line. The first car to finish locks out the data from the second car, and the system can be reset by pressing S1.

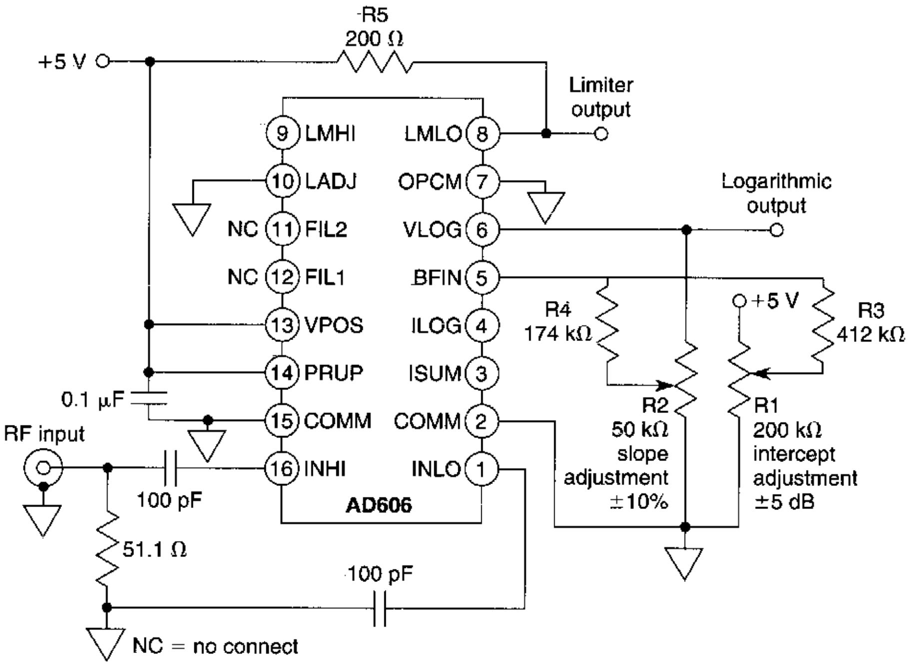
CURRENT INDICATOR



WILLIAM SHEETS *Fig. 39-2*

An LED requires 1.5 to 3 V across its terminals to light. This circuit uses a resistor shunt in series with source and load to produce this drop and cause the LED to light. At higher currents (>100 mA) use limiting resistors in series with LEDs to limit current to a safe value.

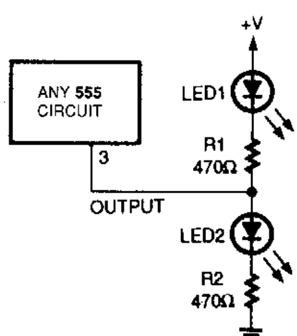
RECEIVER SIGNAL-STRENGTH INDICATOR



ANALOG DEVICES *Fig. 39-3*

Using an AD606 log amplifier, this indicator gives a logarithmic output of +0.3 V at –80-dBm input to +3.5 V at 10-dBm input. Frequency range is to 50 MHz for this IC device.

LED OUTPUT INDICATOR FOR 555 CIRCUITS



POPULAR ELECTRONICS, JANUARY 1994, P. 73 *Fig. 39-4*

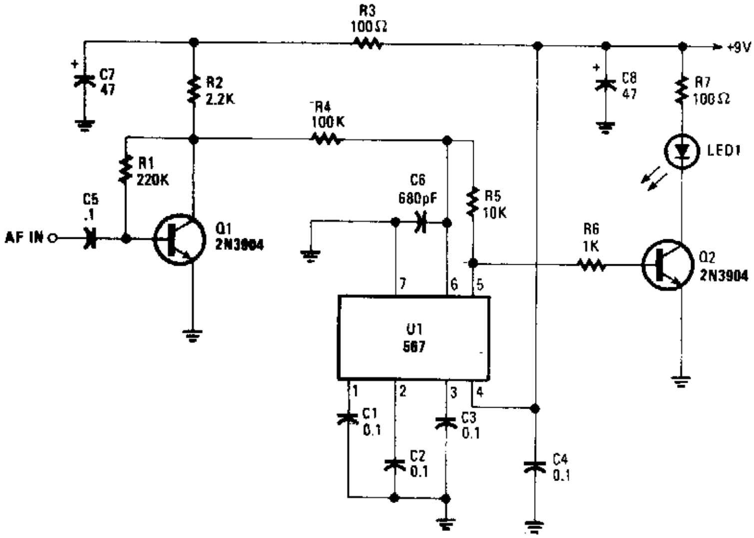
A pair of LEDs connected as shown here can be used with just about any low-frequency 555 oscillator to give high-/low-output indications. When the output goes high LED2 turns on, and when the output goes low LED1 turns on.

40

Infrared Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

AUDIO-MODULATED IR TRANSMITTER

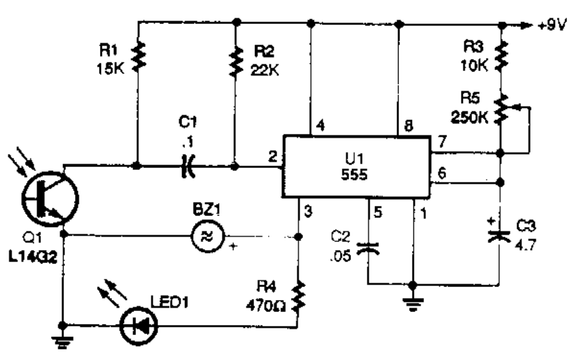


POPULAR ELECTRONICS

Fig. 40-1

This circuit produces an effect similar to frequency modulation (FM) by varying the voltage at pin 6 of the PLL using an audio signal. The FM IR signal can be picked up by a receiver with an FM detector suitably tuned.

AUDIBLE IR DETECTOR



POPULAR ELECTRONICS

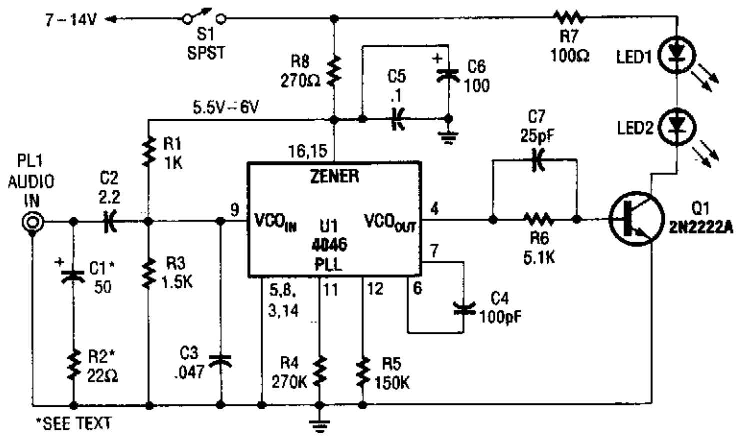
Fig. 40-2

An IR-detector circuit offers an audible (as well as a visual) output, and also stretches the on time of the detected pulse to make the output easier to see, as shown.

Photoresistor Q1 detects a remote's IR output pulse and sends a negative-going pulse to the trigger input (pin) of the 555 IC, U1. The 555 is connected in a one-shot timer circuit; the output (pin 3) on time is set by the values of C_3, R_3, and R_5. When an input pulse is detected, pin 3 goes high, lighting LED1 and activating the piezo buzzer, BZ1.

For longer output pulses, set R_5 to its maximum resistance value. To lengthen the circuit's on-time range, increase the value of C_3, and to shorten the on-time range lower the value of C_3.

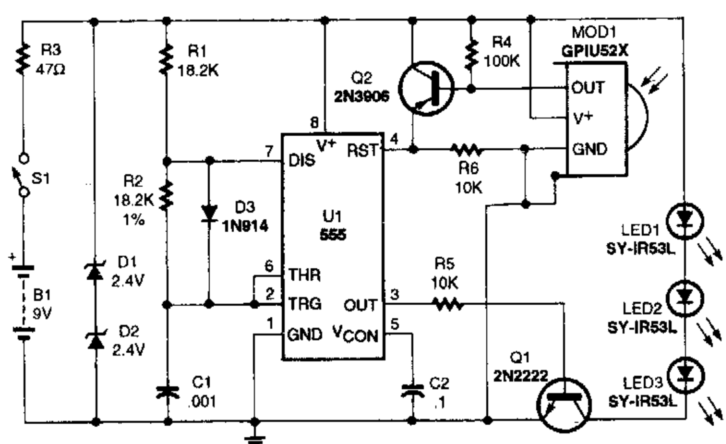
WIRELESS IR HEADPHONE TRANSMITTER



POPULAR ELECTRONICS *Fig. 40-3*

Audio input from PL1 frequency modulates the VCO section of a 4046 PLL chip. The VCO output drives Q1, a switching transistor. Q1 drives two IR LEDs. The signal produced is around 100 kHz, FM carrier VCO sensitivity is around 7.5 kHz/V.

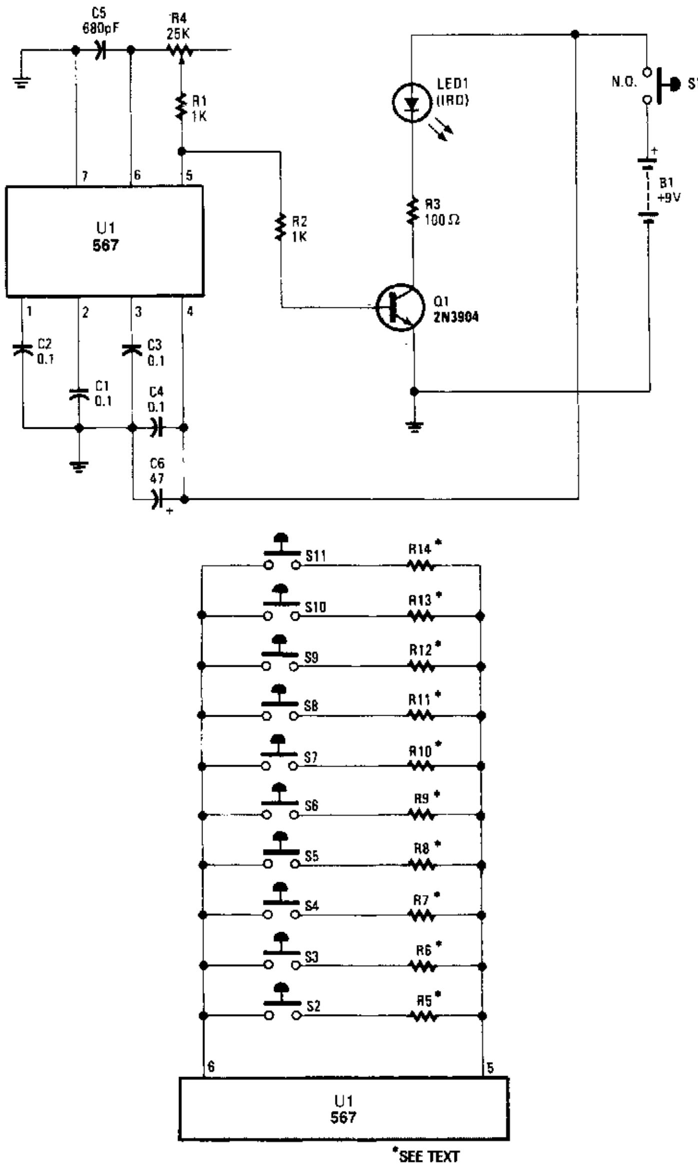
TV REMOTE-CONTROL RELAY



POPULAR ELECTRONICS *Fig. 40-4*

This circuit functions as an IR "repeater" to extend the range of your TV remote control. MOD1 is a P/N GP1U52X IR detector and the receiver is available as Radio Shack P/N 276-137.

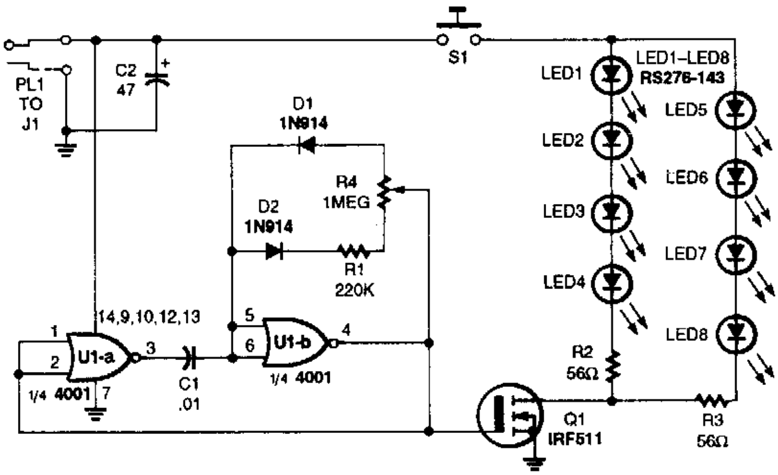
SINGLE-TONE INFRARED CONTROL TRANSMITTER



POPULAR ELECTRONICS ***Fig. 40-5***

A modulated beam of IR light is produced by this transmitter. This circuit can be used for on/off controls or tone (CW) communications. The pot can be replaced by several pushbuttons and resistors, as shown for multitone applications.

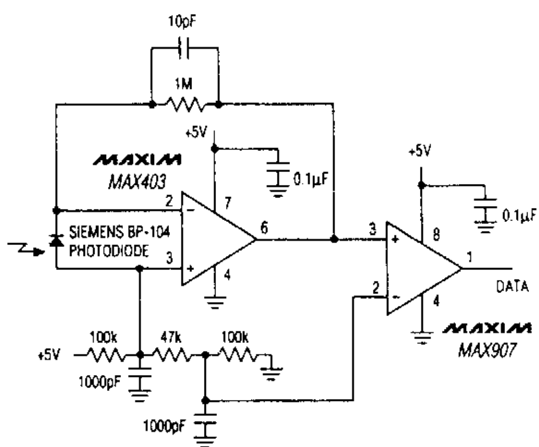
IR ILLUMINATOR FOR NIGHT-VISION TV CAMERAS AND SCOPES



POPULAR ELECTRONICS *Fig. 40-6*

This source uses LEDs and an astable oscillator to control the switch, duty-cycle, and effective IR illumination output.

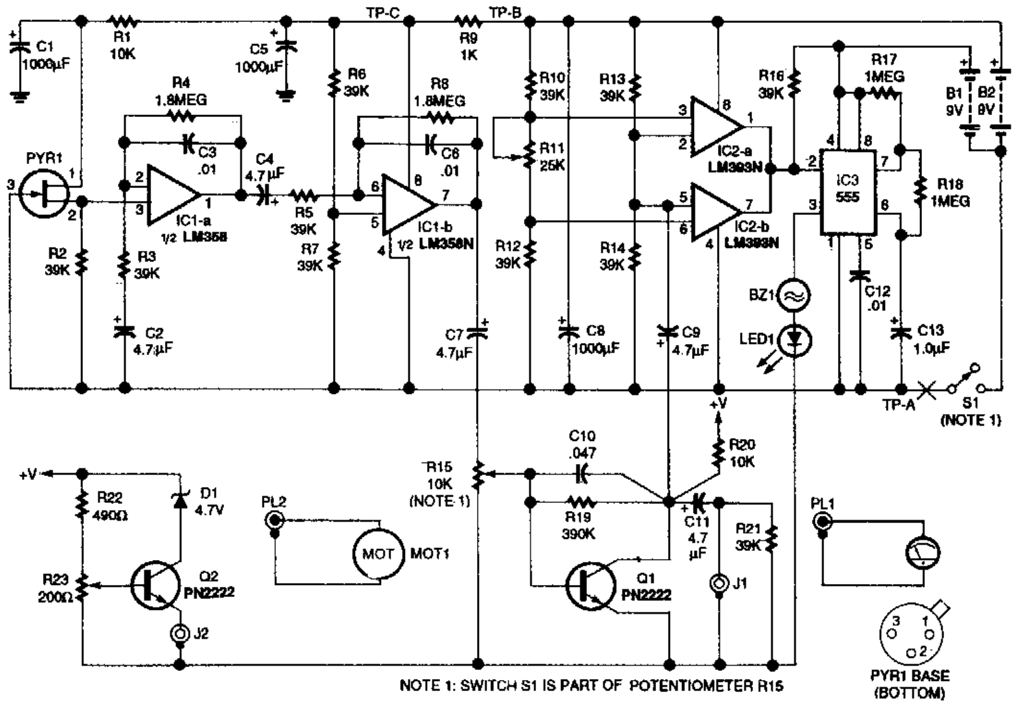
LOW-POWER INFRARED DATA-LINK RECEIVER



MAXIM *Fig. 40-7*

The Maxim MAX403 in this circuit consumes only 1 mA and is capable of speeds over 1 MBPS.

INFRARED BODY-HEAT DETECTOR

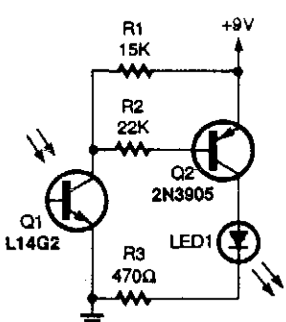


ELECTRONICS NOW

Fig. 40-8

This circuit uses a pyroelectric detector to detect IR emissions in the 6- to 14-micron range. It is useful for security or infrared experiments. PYR1 is a pyroelectric IR detector. The unit should be mounted in a case with an IR lens to focus energy on the detector.

IR DETECTOR CIRCUIT

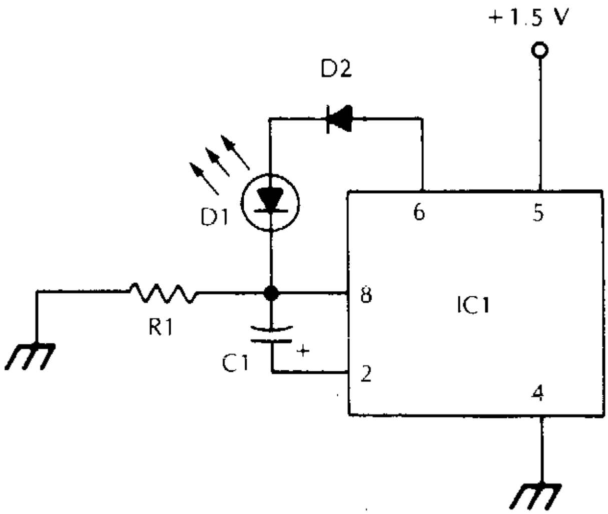


The circuit uses an IR phototransistor, Q1, to detect a remote control's IR output signal. A PNP transistor, Q2, then amplifies Q1's output and lights LED1. That indicates that an infrared signal has been detected by the phototransistor, or in other words, that your remote control works.

POPULAR ELECTRONICS

Fig. 40-9

STEADY-TONE INFRARED TRANSMITTER

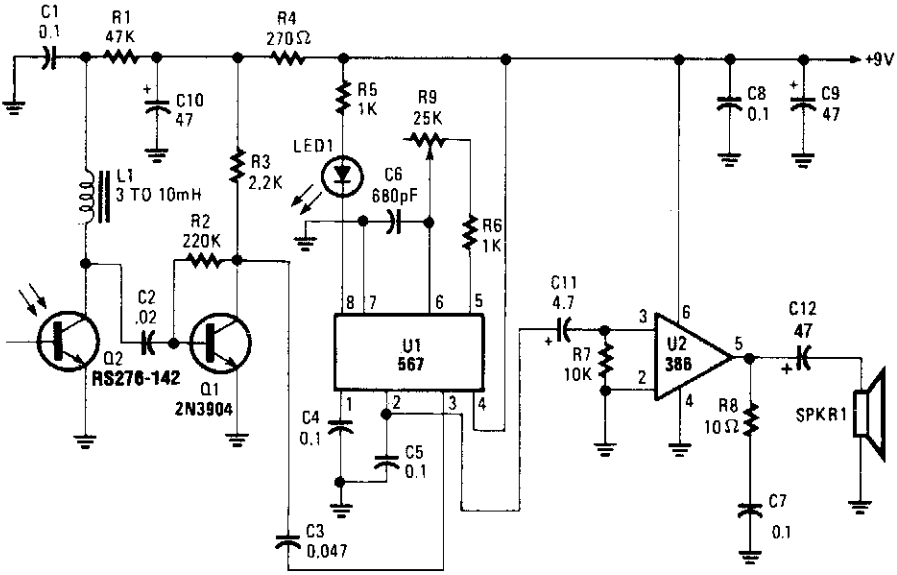


IC1	LM3909 LED flasher/oscillator IC
D1	infrared LED
D2	diode (1N4148, 1N914, or similar)
C1	1 μF 5 V electrolytic capacitor
R1	1.5 kΩ ¼ W 5% resistor

McGRAW-HILL *Fig. 40-10*

This oscillator pulses an IR LED at about 1000 Hz. It should be useful as a test for lining up IR communications links or setting up fiber-optic cables, etc.

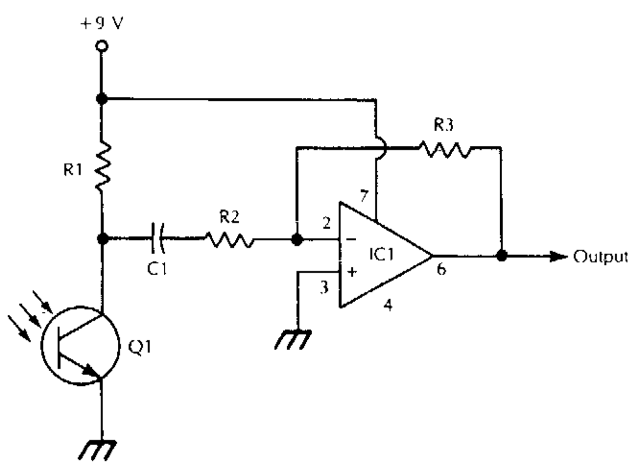
FM INFRARED RECEIVER FOR AUDIO RECEPTION



POPULAR ELECTRONICS *Fig. 40-11*

Modulated IR energy strikes Q2, a phototransistor. Q1 is a tuned amplifier, and feeds PLL detector U1. U2 is an audio amplifier that drives a speaker.

GENERAL-PURPOSE IR RECEIVER

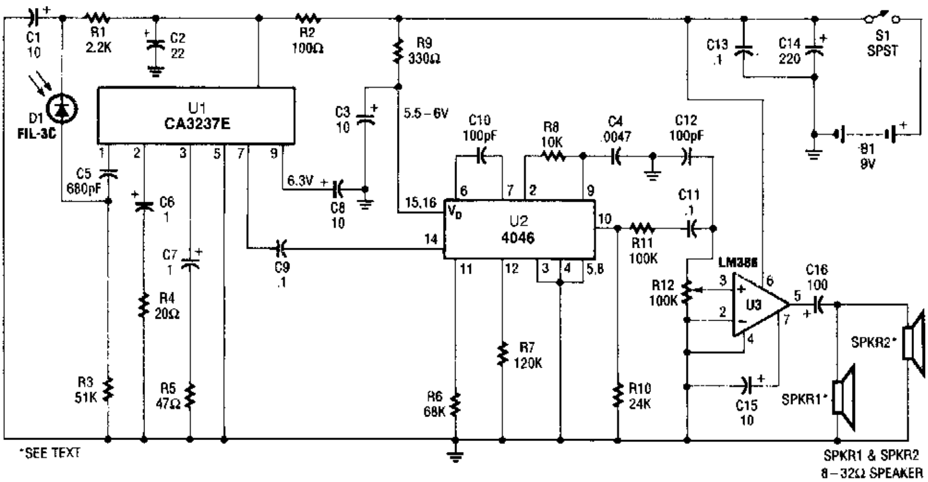


IC1	741 op amp
Q1	infrared phototransistor
C1	0.01 µF capacitor
R1	100 kΩ ¼ W 5% resistor
R2	2.2 kΩ ¼ W 5% resistor
R3	1 MΩ ¼ W 5% resistor

McGRAW-HILL *Fig. 40-12*

Suitable for amplitude-modulated IR beams, this receiver provides an audio signal that corresponds to the modulation envelope. Phototransistor Q1 should be properly mounted and shielded from stray light. This receiver should drive a small earphone directly.

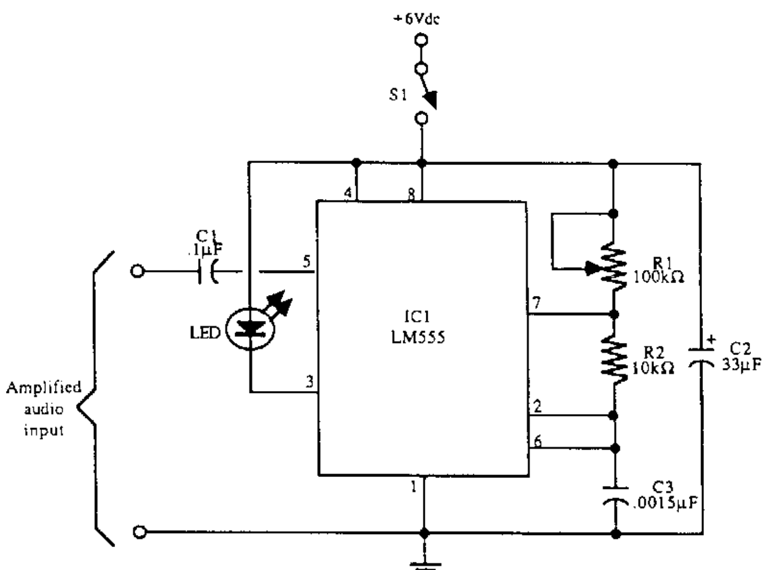
WIRELESS IR HEADPHONE RECEIVER



POPULAR ELECTRONICS ***Fig. 40-13***

A photodiode D1 feeds high gain IR remote control preamp IC, a CA3237E. U2 is a PLL FM detector tuned to around 100 kHz. The detector output is amplified by U3 and it can drive a speaker or a set of headphones.

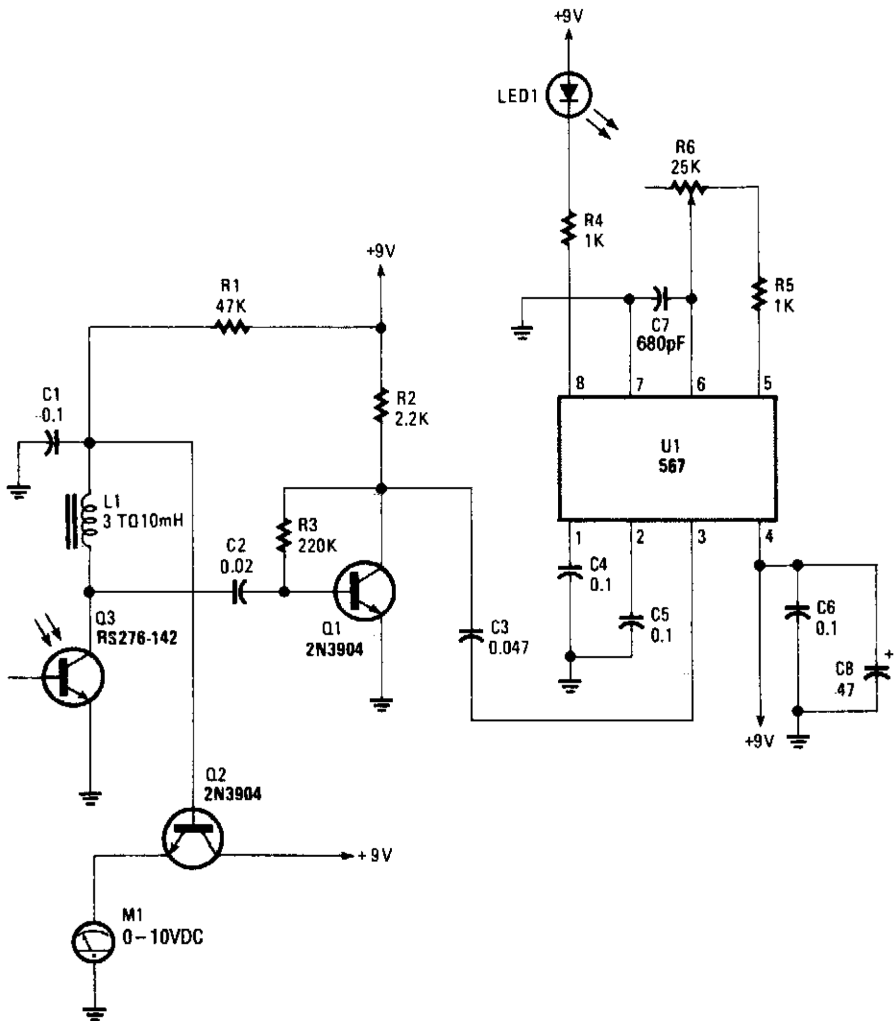
PULSE FREQUENCY-MODULATED IR TRANSMITTER



McGRAW-HILL ***Fig. 40-14***

Schematic diagram for the pulse frequency-modulated LED transmitter. Adjust the frequency by rotating R1. With components shown, the frequency range is between 8 and 48 kHz.

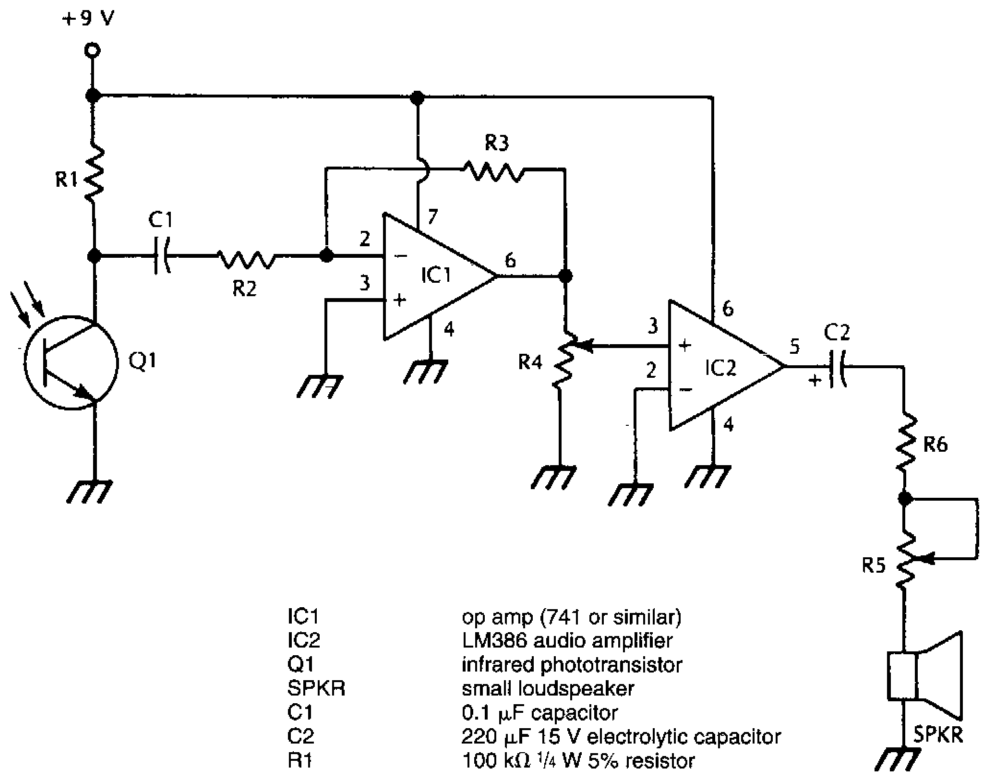
SINGLE-TONE INFRARED RECEIVER



POPULAR ELECTRONICS *Fig. 40-15*

Phototransistor Q3 acts as a sensor that detects modulated IR energy. Q1 is an amplifier and U1 is a tone decoder. LED1 lights on reception of an IR signal with proper tone modulation.

AUDIBLE-OUTPUT INFRARED RECEIVER



IC1	op amp (741 or similar)
IC2	LM386 audio amplifier
Q1	infrared phototransistor
SPKR	small loudspeaker
C1	0.1 μF capacitor
C2	220 μF 15 V electrolytic capacitor
R1	100 kΩ 1/4 W 5% resistor
R2	1 kΩ 1/4 W 5% resistor
R3	680 kΩ 1/4 W 5% resistor
R4	25 kΩ potentiometer
R5	500 Ω potentiometer
R6	47 Ω 1/4 W 5% resistor

McGRAW-HILL ***Fig. 40-16***

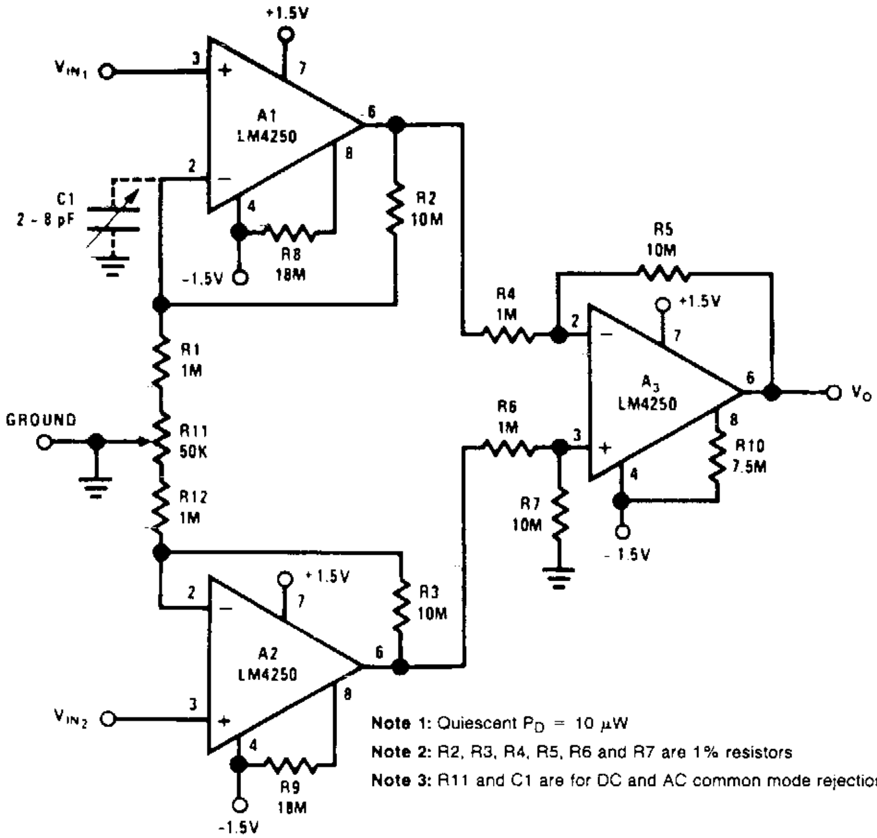
This receiver is designed to demodulate amplitude-modulated (AM) IR light beams and will drive a loudspeaker. R5 is an auxiliary volume control and it could be omitted. Q1 should be suitably mounted and shielded from stray light pickup.

41

Instrumentation Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

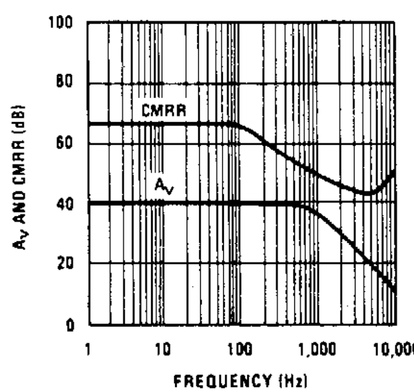
×100 INSTRUMENTATION AMPLIFIER



Note 1: Quiescent P_D = 10 μW

Note 2: R2, R3, R4, R5, R6 and R7 are 1% resistors

Note 3: R11 and C1 are for DC and AC common mode rejection adjustments

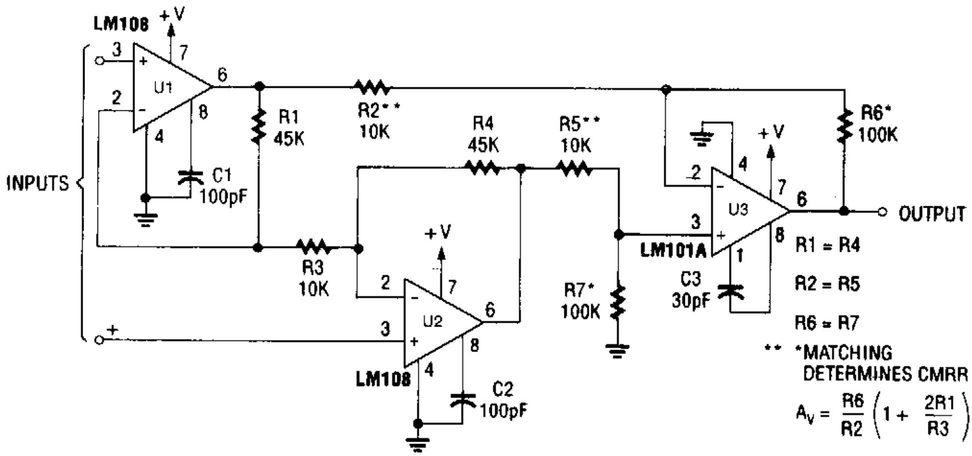


NATIONAL SEMICONDUCTOR

Fig. 41-1

CMRR vs. frequency.

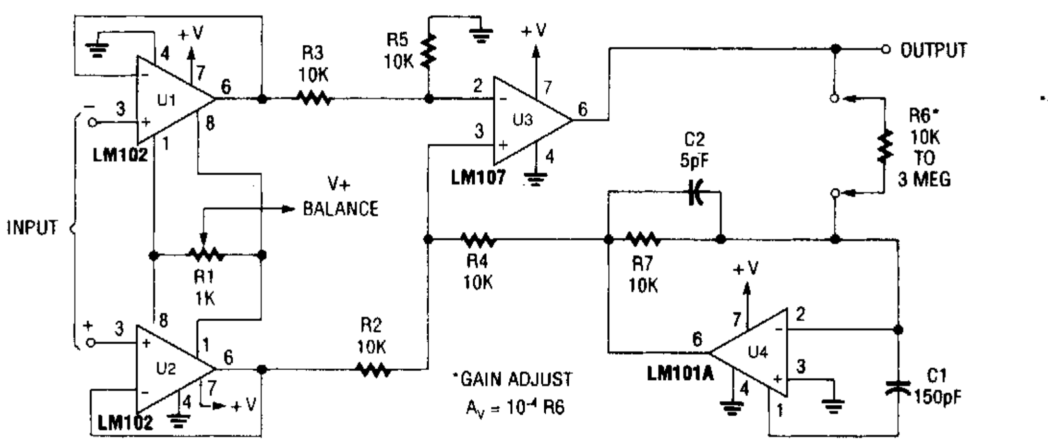
INSTRUMENTATION AMPLIFIER



LINEAR TECHNOLOGY

Fig. 41-2

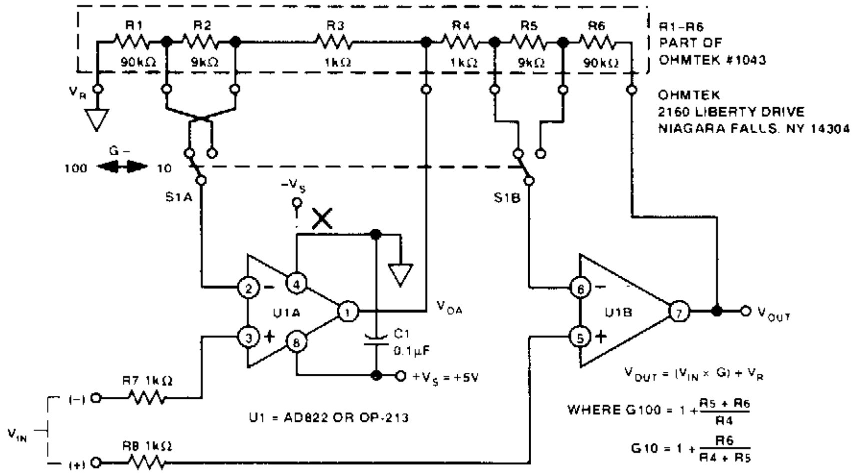
VARIABLE-GAIN DIFFERENTIAL INPUT INSTRUMENTATION AMPLIFIER



POPULAR ELECTRONICS

Fig. 41-3

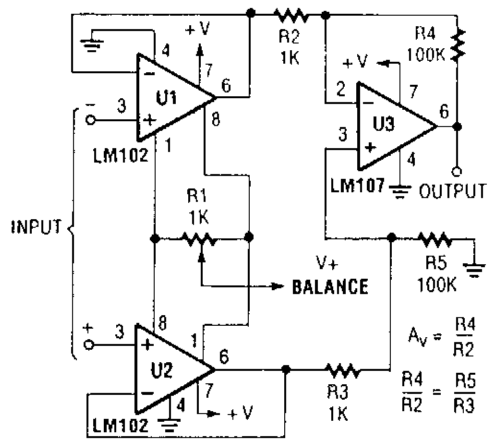
PROGRAMMABLE GAIN INSTRUMENTATION AMPLIFIER FOR SINGLE-SUPPLY APPLICATIONS



ANALOG DEVICES *Fig. 41-4*

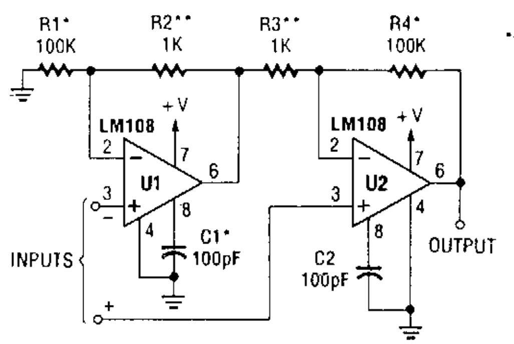
This is a two-op-amp programmable-gain instrumentation amplifier for single-supply applications. U1A and U1B are Analog Devices AD822 or OP-213 ICs.

DIFFERENTIAL-INPUT INSTRUMENTATION AMPLIFIER



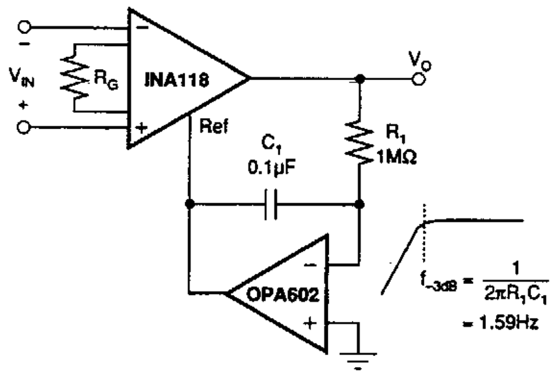
POPULAR ELECTRONICS *Fig. 41-5*

HIGH INPUT-IMPEDANCE INSTRUMENTATION AMPLIFIER



POPULAR ELECTRONICS *Fig. 41-6*

ac-COUPLED INSTRUMENTATION AMPLIFIER

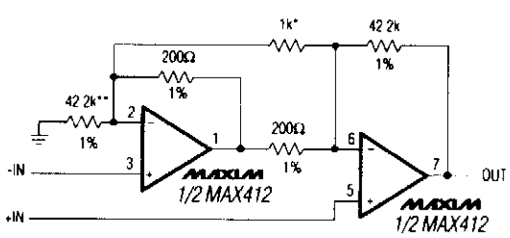


DESIRED GAIN	R_G (Ω)	NEAREST 1% R_G (Ω)
1	NC	NC
2	50.00k	49.9k
5	12.50k	12.4k
10	5.556k	5.62k
20	2.632k	2.61k
50	1.02k	1.02k
100	505.1	511
200	251.3	249
500	100.2	100
1000	50.05	49.9
2000	25.01	24.9
5000	10.00	10
10000	5.001	4.99

BURR-BROWN

Fig. 41-7

LOW-NOISE INSTRUMENTATION AMPLIFIER



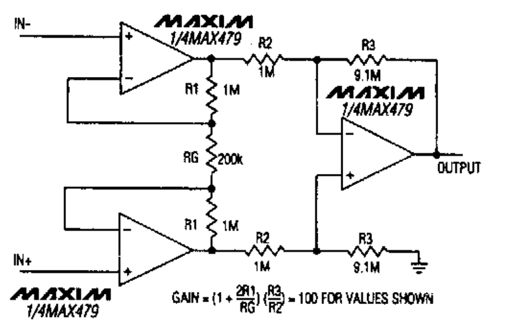
* TRIM FOR GAIN
** TRIM FOR COMMON-MODE REJECTION

MAXIM

Fig. 41-8

A Maxim MAX412 IC amplifier is used in this circuit. The supply current is ±5 V at 5 mA.

LOW-POWER INSTRUMENTATION AMPLIFIER

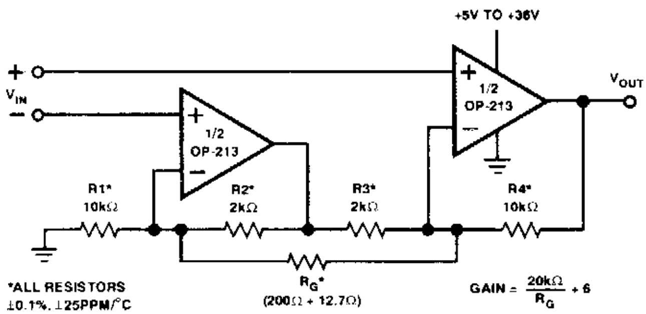


MAXIM

Fig. 41-9

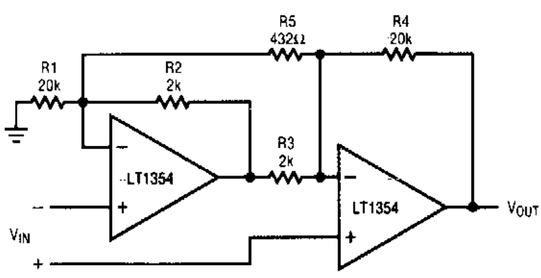
This amplifier requires less than 20 mA from a ±15-V supply.

ULTRA-LOW-NOISE SINGLE-SUPPLY INSTRUMENTATION AMPLIFIER



ANALOG DEVICES *Fig. 41-10*

INSTRUMENTATION AMPLIFIER



$$A_V = \frac{R4}{R3}\left[1 + \frac{1}{2}\left(\frac{R2}{R1} + \frac{R3}{R4}\right) + \frac{R2+R3}{R5}\right] = 104$$

TRIM R5 FOR GAIN
TRIM R1 FOR COMMON-MODE REJECTION
BW = 120kHz

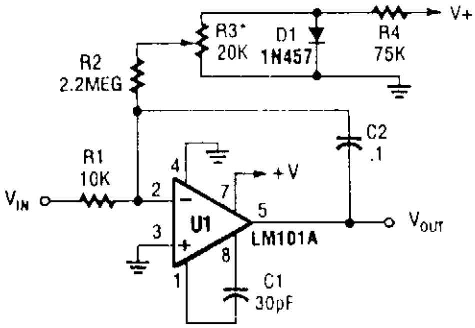
LINEAR TECHNOLOGY *Fig. 41-11*

42

Integrator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

INTEGRATOR WITH BIAS-CURRENT COMPENSATION

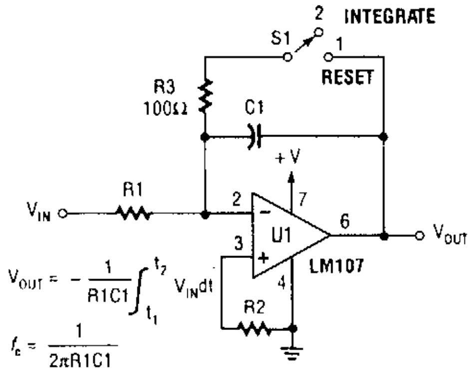


*ADJUST FOR ZERO INTEGRATOR DRIFT. CURRENT DRIFT TYPICALLY 0.1 n/A°C OVER −55°C TO 125°C TEMPERATURE RANGE.

POPULAR ELECTRONICS

Fig. 42-1

SIMPLE INTEGRATOR



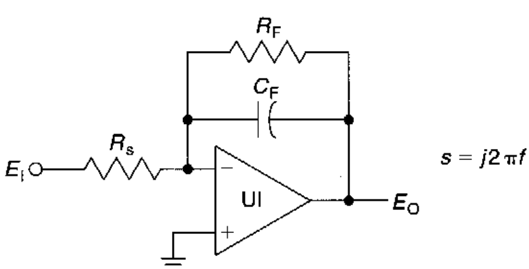
R1 = R2

FOR MINIMUM OFFSET ERROR DUE TO INPUT BIAS CURRENT

POPULAR ELECTRONICS

Fig. 42-2

ac INTEGRATOR



$$\frac{E_O}{E_I}(s) = \frac{R_F}{R_S(1 + sR_FC_F)} \approx \frac{1}{R_FC_F} \text{ for } s \gg \frac{1}{R_FC_F}$$

WILLIAM SHEETS

Fig. 42-3

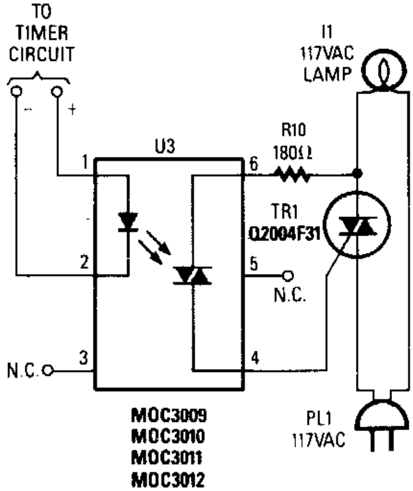
This op-amp circuit can be used with a wide variety of op amps. The values of R_f and R_i depend on gain, but will be 1 kΩ to 1 MΩ in most cases. C_f depends on the pole frequency needed. U1 is a 741-type op amp, etc.

43

Interface Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

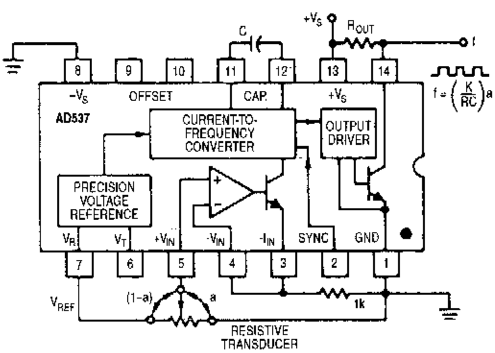
TIMER/ac LINE INTERFACE



POPULAR ELECTRONICS *Fig. 43-1*

This circuit illustrates the use of an optoisolator to enable the control of a triac connected to the ac line and load, while maintaining dc and ac isolation between the ac line and the timer circuit. A 555 or other timer circuit can be used.

INTERFACING RESISTIVE TRANSDUCERS



ANALOG DEVICES *Fig. 43-2*

All types of resistive-element transducers, such as servo-pots, level indicators, thermistors, photosensors, strain gages, and so on, can be directly connected to the AD537. The scale-correction factor, K, is a function of resistance, varying from 0.65 to 0.98 for values from 3 to 100 kΩ.

44

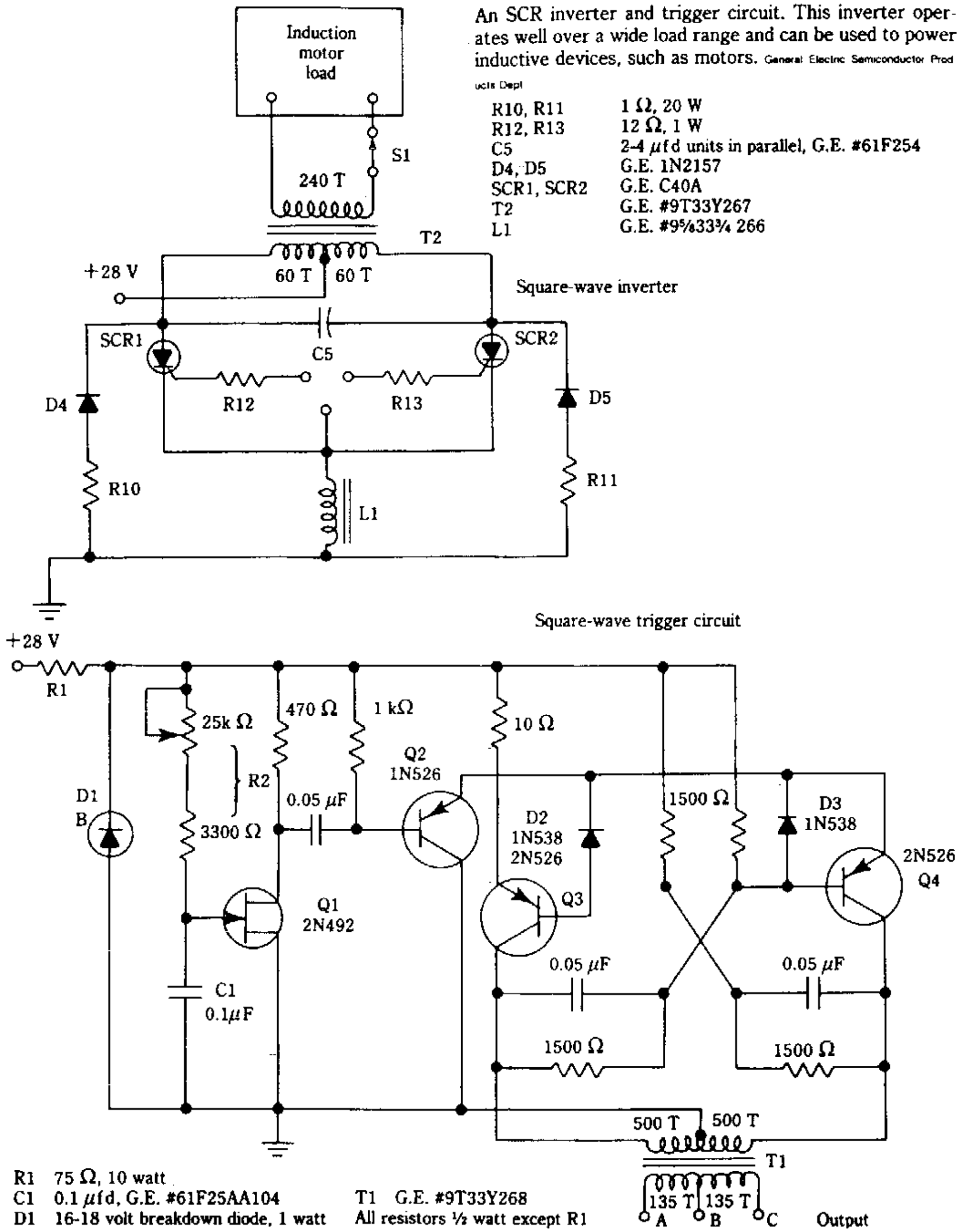
Inverter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SCR INVERTER AND TRIGGER CIRCUIT

An SCR inverter and trigger circuit. This inverter operates well over a wide load range and can be used to power inductive devices, such as motors. General Electric Semiconductor Products Dept

R10, R11	1 Ω, 20 W
R12, R13	12 Ω, 1 W
C5	2-4 μfd units in parallel, G.E. #61F254
D4, D5	G.E. 1N2157
SCR1, SCR2	G.E. C40A
T2	G.E. #9T33Y267
L1	G.E. #9⅝33¾ 266



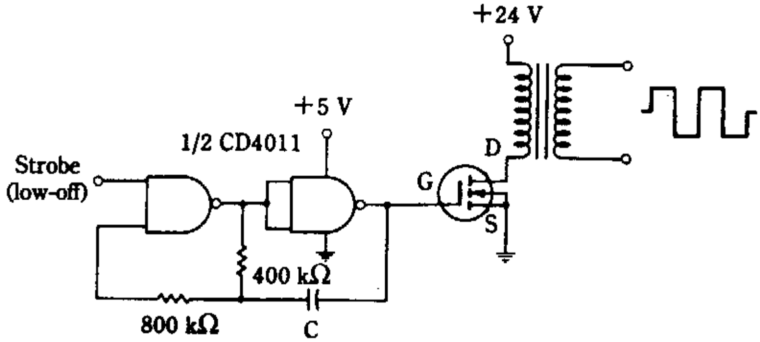
R1 75 Ω, 10 watt
C1 0.1 μfd, G.E. #61F25AA104
D1 16-18 volt breakdown diode, 1 watt

T1 G.E. #9T33Y268
All resistors ½ watt except R1

McGRAW-HILL *Fig. 44-1*

In this circuit, L1 and C5 are used as commutating elements. L1 resonates with C5 at the frequency corresponding to the half period of the waveform.

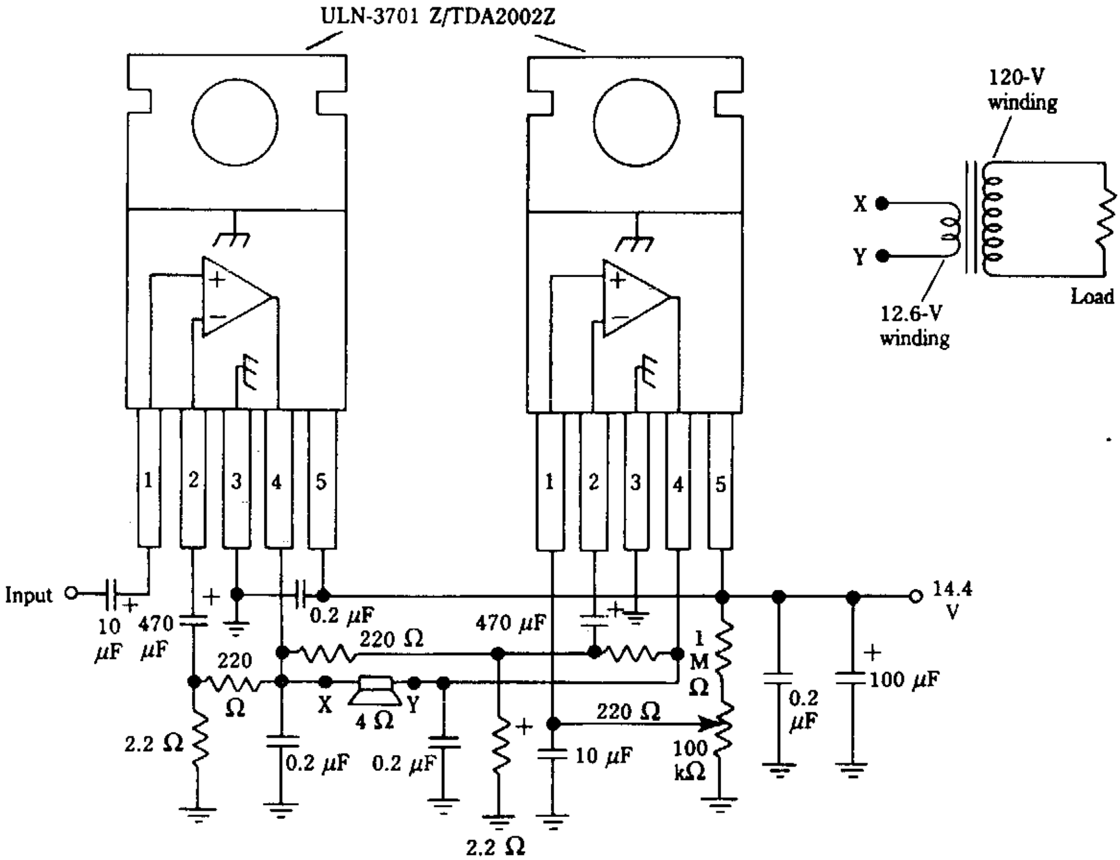
SIMPLE INVERTER



McGRAW-HILL

Fig. 44-2

VEHICLE AUDIO AMPLIFIER INVERTER

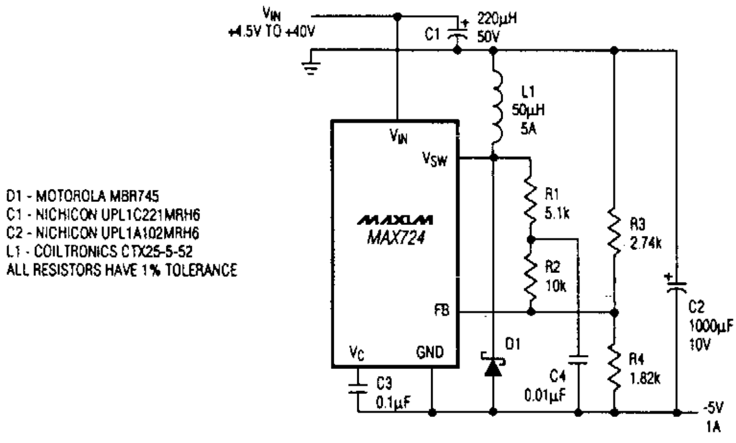


McGRAW-HILL

Fig. 44-3

An audio amplifier can drive a step-up transformer to obtain 120 Vac.

POSITIVE-TO-NEGATIVE dc/dc INVERTER

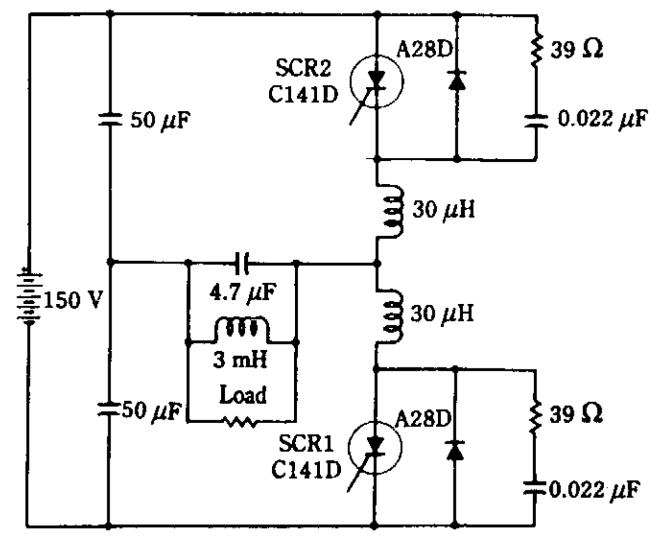


MAXIM

Fig. 44-4

If a source of negative 5 Vdc is needed and only a positive supply is available, this circuit can be used.

1-kW 10-kHz SINE-WAVE INVERTER



McGRAW-HILL

Fig. 44-5

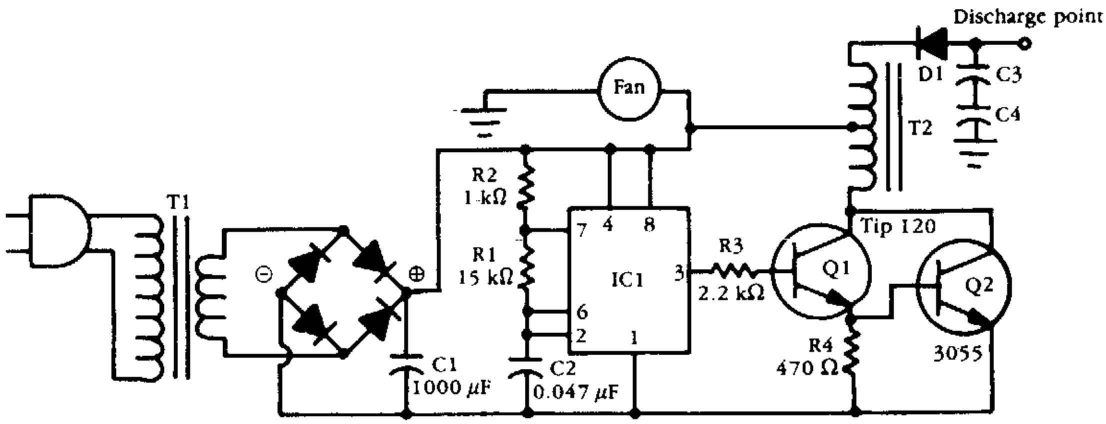
SCRs can produce considerable power at frequencies up to 30 kHz or more. This circuit can supply 1 kW at 10 kHz. The load is shown as an equivalent load, and practically this will be the primary of the transformer for isolation purposes. The power supply can be a 120-V bridge rectifier and filter combination.

45

Ion Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

NEGATIVE ION GENERATOR

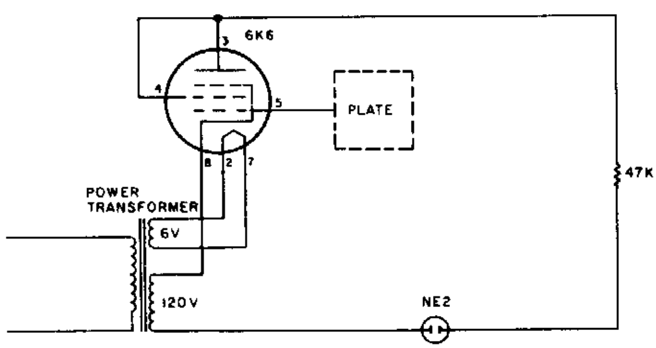


McGRAW-HILL

Fig. 45-1

An NE555 drives a Darlington connected pair of transistors. T1 is a small high-voltage transformer or auto ignition coil, B/W TV flyback, etc. C3, C4, and D1 must be rated for 10 to 15 kV. The fan blows air across the discharge point.

ION-SENSING ELECTROSCOPE

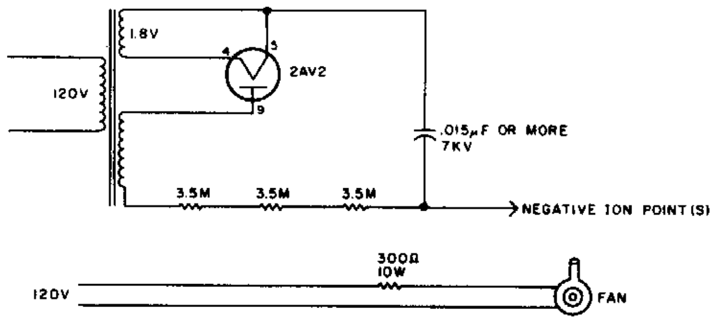


73 AMATEUR RADIO TODAY

Fig. 45-2

Negative ions are sensed by a plate antenna. A negative charge induced on the plate cuts off a vacuum tube, causing the neon indicator to go out.

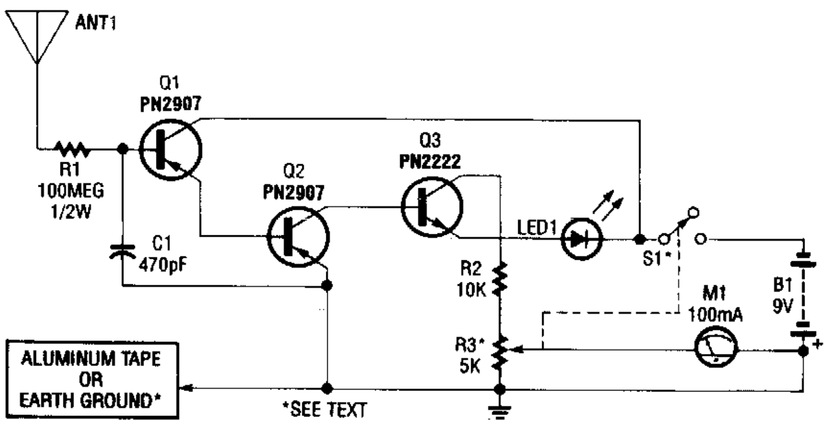
NEGATIVE ION GENERATOR



73 AMATEUR RADIO TODAY ***Fig. 45-3***

In this circuit, air is circulated past a pointed electrode that has a high negative voltage applied to it. The transformer is a small 4- to 6-kV output type with a filament winding. A good source of parts is a discarded electronic bug catcher.

ION DETECTOR



ELECTRONIC HOBBYISTS HANDBOOK ***Fig. 45-4***

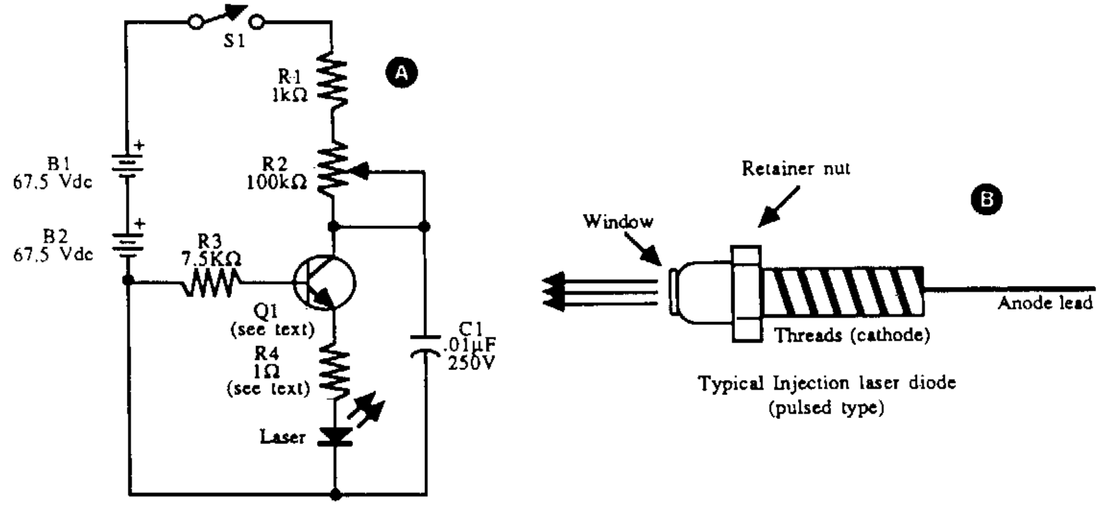
This circuit detects static charges and free ions in the air. It can be used to indicate the presence of ion emissions, high-voltage leakage, static electricity, electrostatic fields, etc. The ground connection is made by either an earth ground or by touching the aluminum foil electrode with your hand. M1 is a 100-μA meter. R3 is a sensitivity control.

46

Laser Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

HIGH-CURRENT DRIVE CIRCUIT FOR SINGLE HETEROSTRUCTURE LASER DIODES



(A) High-current drive circuit for a single heterostructure laser diode. (B) Power leads for the typical sh laser diode, showing single lead for the anode.

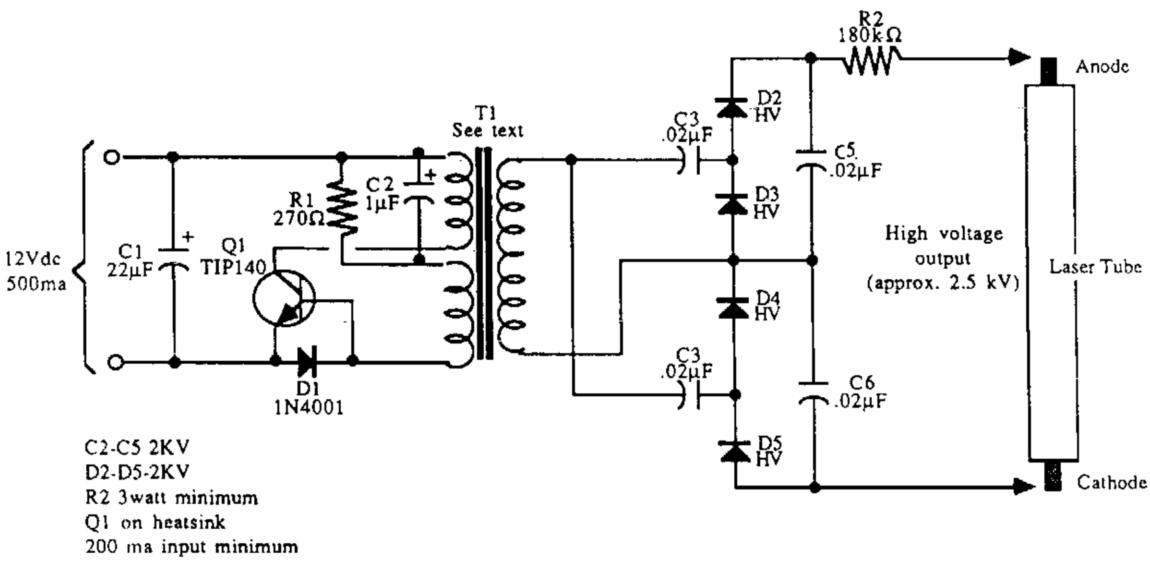
R1	1 kilohm resistor
R2	100 kilohm potentiometer
R3	7.5 kilohm resistor
R4	1 ohm resistor, carbon composition, 5 watts
C1	0.01 μF capacitor, 250 V or higher
Q1	2N2222 or equivalent; see text
B1,B2	67.5 Vdc batteries
Misc.	Single heterostructure laser diode, heatsink

All resistors are 5 to 10 percent tolerance, ¼ watt, unless otherwise indicated.

McGRAW-HILL **Fig. 46-1**

The transistor is operated in the avalanche mode. You might need to try several 2N2222 devices before finding one that oscillates. R2 is adjusted for optimum oscillation. This supply provides pulse of 10 to 20 amps at about 50 ns.

12-V HIGH-VOLTAGE SUPPLY FOR HE-NE LASER



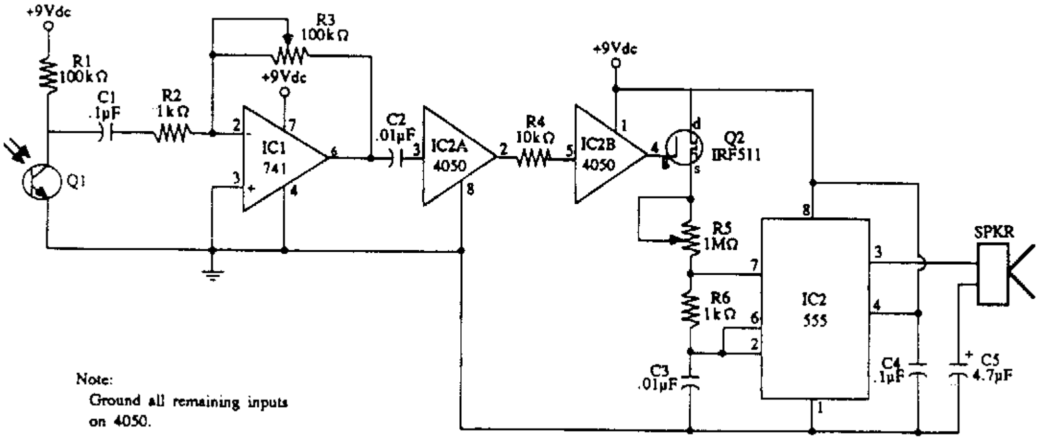
McGRAW-HILL *Fig. 46-2*

T1 is a 6-V:330-V dc/dc inverter transformer with a 57.4:1 turn ratio, rated at 7 W.

R1	270-Ω resistor
R2	180-kΩ resistor, 3 to 5 W
C1	22-μF electrolytic capacitor
C2	1-μF electrolytic capacitor
C3-C6	0.02-μF capacitor, 1 kV or more
D1	1N4001 diode
D2-D5	High-voltage diode (3 kV or more)
Q1	TIP 140 power transistor
T1	High-voltage dc-to-dc converter transformer; see text for specifications

All resistors are 5 to 10% tolerance, ¼ W, unless otherwise indicated. All capacitors are 10 to 20% tolerance, rated 35 V or more, unless otherwise indicated.

LIGHT-BEAM RECEIVER AND SOUND EFFECTS GENERATOR FOR LASER PISTOLS

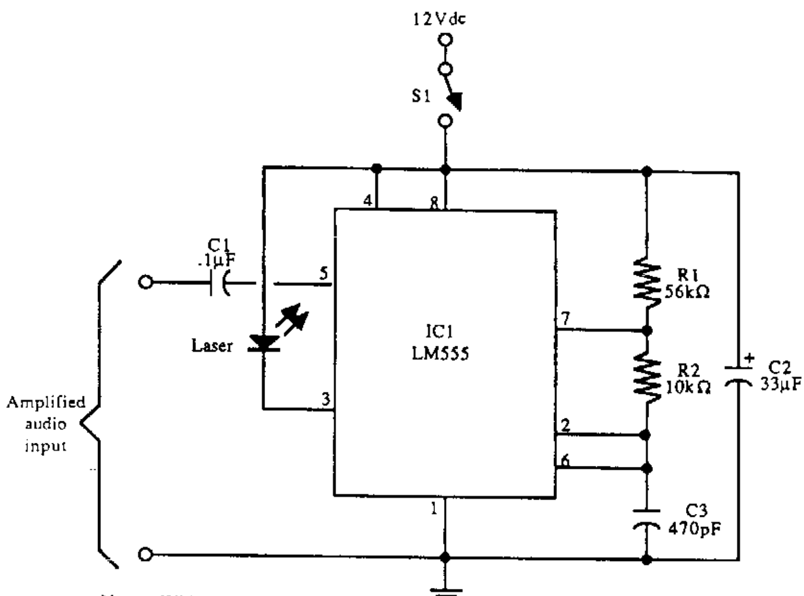


McGRAW-HILL

Fig. 46-3

Schematic diagram for light beam amplifier and sound-effects generator (using a 555 timer IC and speaker). The light striking Q1 generates a siren-like sound.

LASER DIODE TRANSMITTER



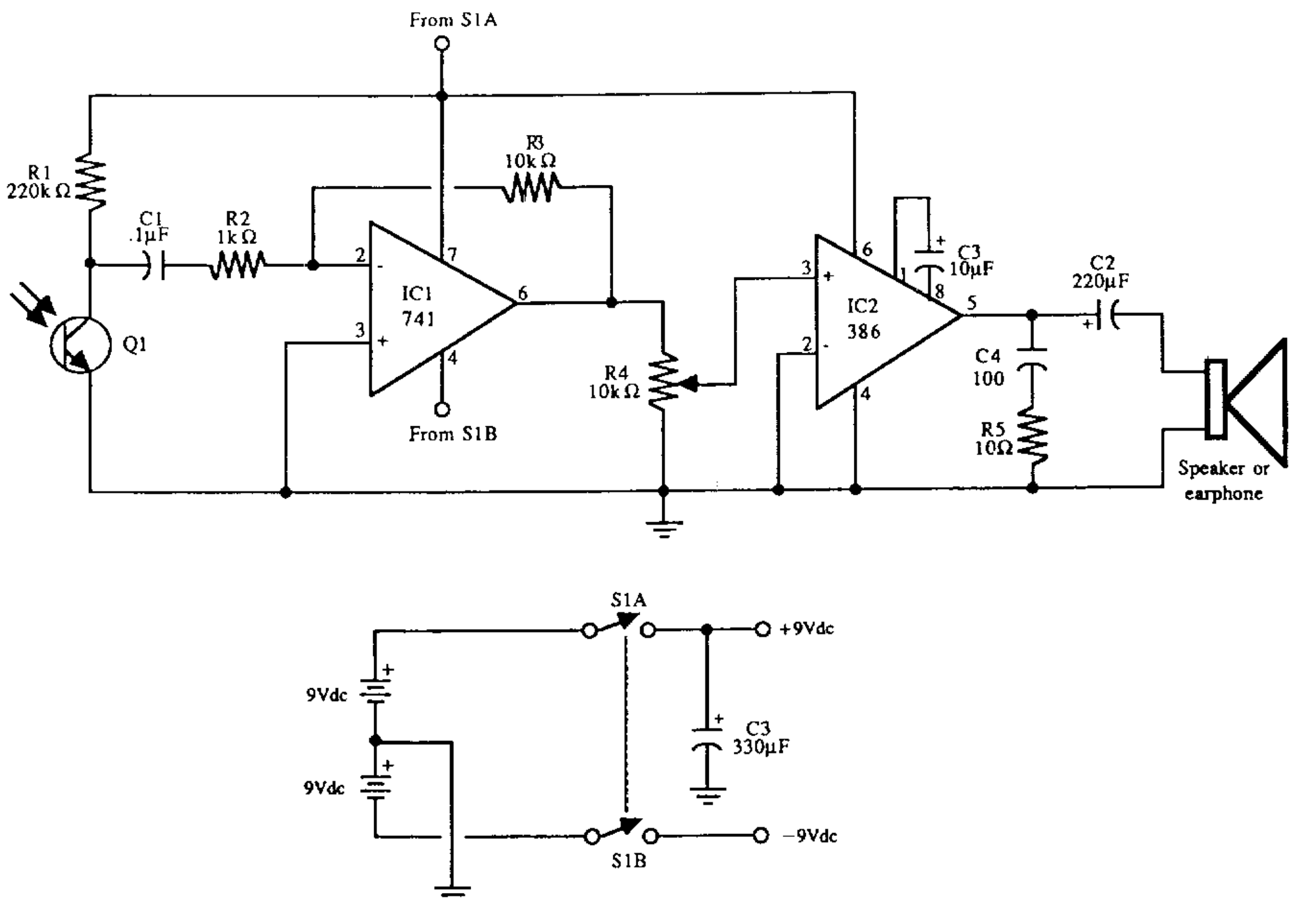
Note: With components show, center frequency is 40.31 kHz; replace R1 with 100K pot to adjust center frequency.

Calculate center frequency using the formula: $f = \frac{1.44}{C3\,(R1 + 2R2)}$

McGRAW-HILL

Fig. 46-4

IR LASER LIGHT DETECTOR

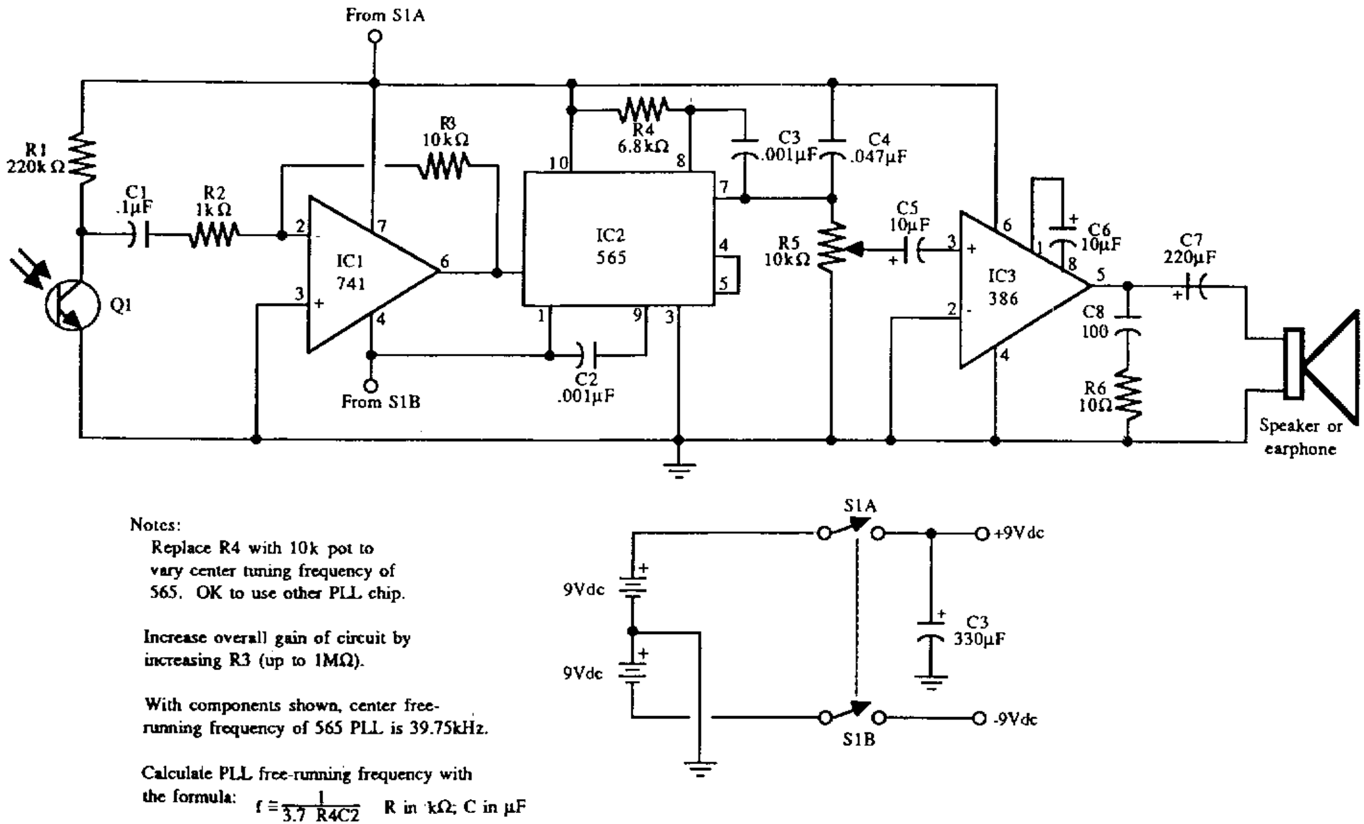


McGRAW-HILL

Fig. 46-5

The universal laser light detector. The output of the LM386 audio amplifier can be connected to a small 8-Ω speaker or earphone. Two 9-V batteries provide power. Decrease R1 to lower sensitivity; increase R3 to increase gain of the op amp (avoid very high gain or the op amp might oscillate). Q1 is an infrared phototransistor.

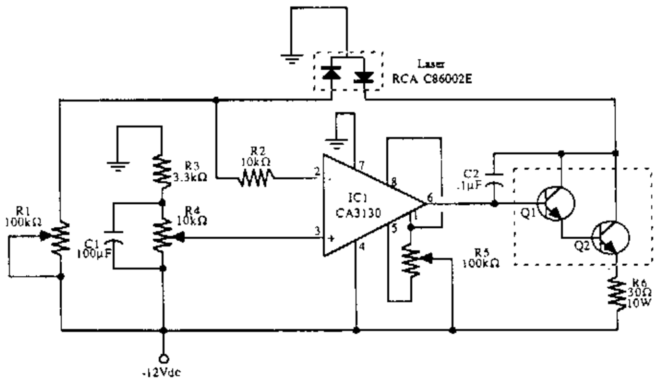
PLL IR LASER LIGHT RECEIVER



McGRAW-HILL *Fig. 46-6*

Circuit schematics for the 555-based PLL laser light PFM receiver. Although R4 is shown as a resistor, you might want to substitute it with a 10-kΩ precision potentiometer so that you can "dial in" the center frequency of the transmitter. Experiment with the value of C_1 for the best high-frequency response. Notice that circuit is functionally identical to the laser light detector/receiver shown in the figure, but with the addition of the 565.

OP-AMP DIODE LASER DRIVER



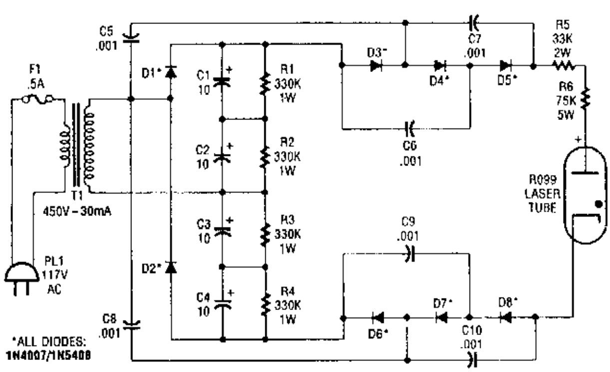
McGRAW-HILL

Fig. 46-7

This circuit is one way to automatically adjust drive current using a discrete op amp. Use the transistors specified or replace them with a suitable Darlington power transistor (such as TIP 120).

IC1	RCA CA 313 operational amplifier
R1, R5	100-kΩ potentiometer
R2	10-kΩ resistor
R3	3.3-kΩ potentiometer
R4	10-kΩ potentiometer
R6	30-Ω, 10-W resistor
C1	100-µF electrolytic capacitor
C2	0.1-µF disc capacitor
Q1	2N2101 transistor
Q2	2N3585 transistor
Laser	RCA C86002 (or equivalent laser diode)

LASER dc SUPPLY

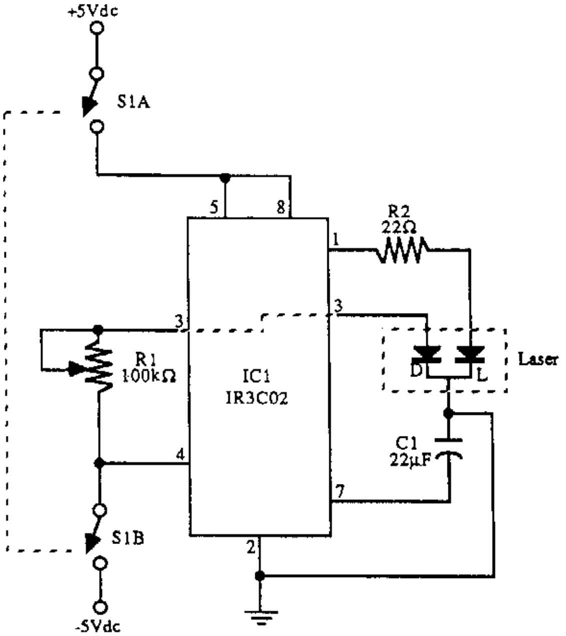


POPULAR ELECTRONICS

Fig. 46-8

The supply provides about 6 kVdc when open circuited, dropping to around 1375 Vdc when loaded. The R099 is a laser tube.

IC LASER DIODE DRIVER

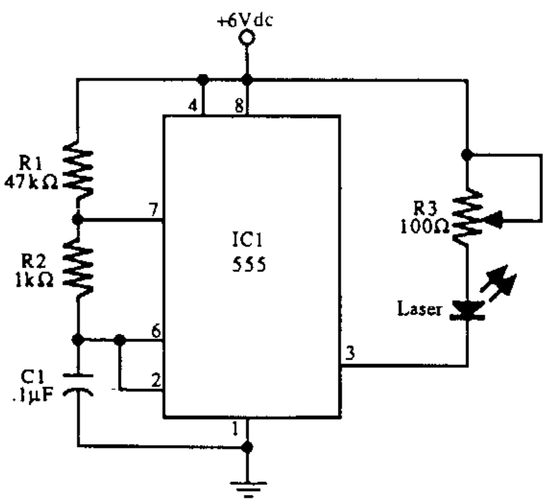


IC	Sharp IR3C02 laser diode driver IC
R1	100-kΩ resistor
R2	22-Ω resistor
C1	22-μF electrolytic capacitor
S1	DPDT switch
Misc.	Double heterostructure laser diode (such as Sharp LT020), heatsink

McGRAW-HILL

Fig. 46-9

PULSED DOUBLE-HETEROSTRUCTURE LASER DRIVER



IC1	555 timer IC
R1	47-kΩ resistor
R2	1-kΩ resistor
R3	100-kΩ potentiometer
C1	0.1-μF disc capacitor
Misc.	Double heterostructure laser diode, heatsink

All resistors are 5 to 10% tolerance, ¼ W. All capacitors are 10 to 20% tolerance, rated 35 V or more.

McGRAW-HILL

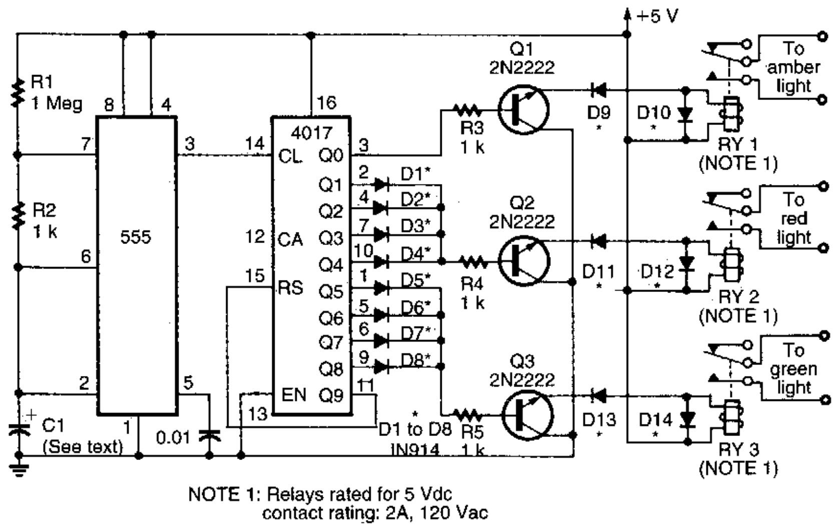
Fig. 46-10

47

Light-Controlled Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

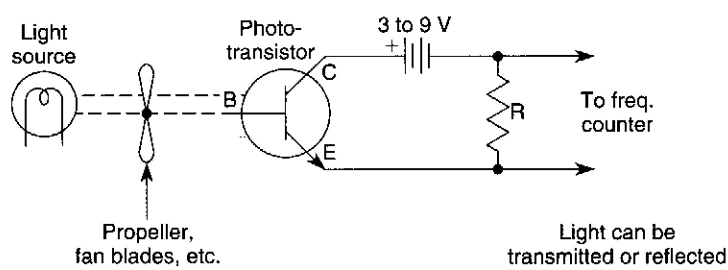
TRAFFIC LIGHT-SEQUENCER CIRCUIT



ELECTRONICS NOW ***Fig. 47-1***

This circuit uses a 555 timer to drive a 4017 counter. The counter outputs drive transistor relay drivers. Time lights "on" can be proportioned by changing connections of outputs of counter.

TACHOMETER ADAPTER



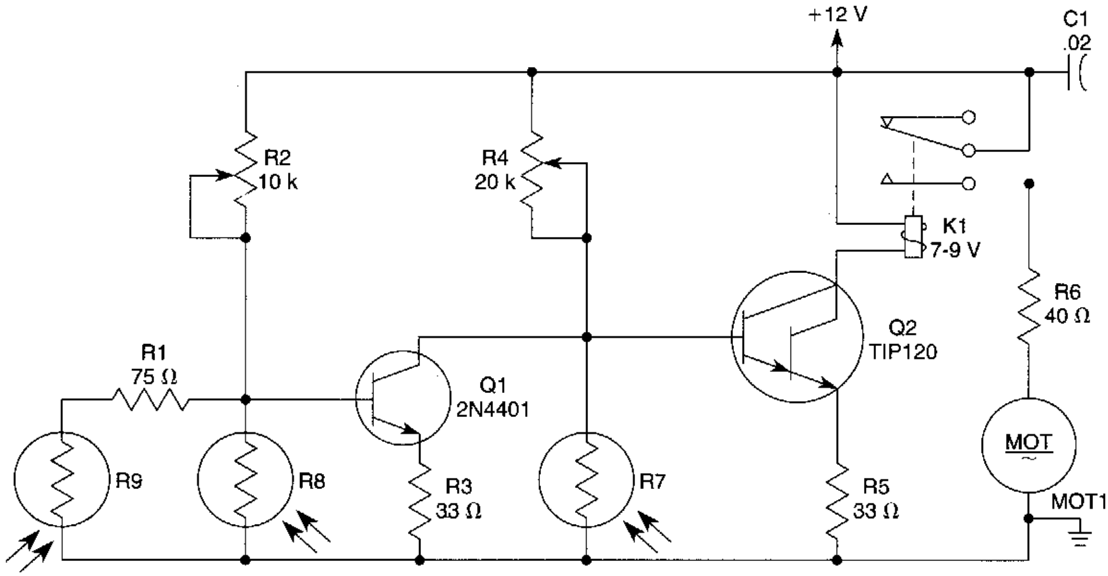
WILLIAM SHEETS ***Fig. 47-2***

Use of a phototransistor and light source can enable a frequency counter to act as a tachometer:

$$RPM = 60 \times \frac{\textit{Frequency Counter Reading Hz}}{\textit{\# Blades or Spokes}}$$

The light source is interrupted by the number of propeller blades, fan blades, spokes, or other marking. R can be anywhere from 1 to 100 kΩ. Try several values for best results.

SUN-TRACKING CIRCUIT FOR SOLAR ARRAYS



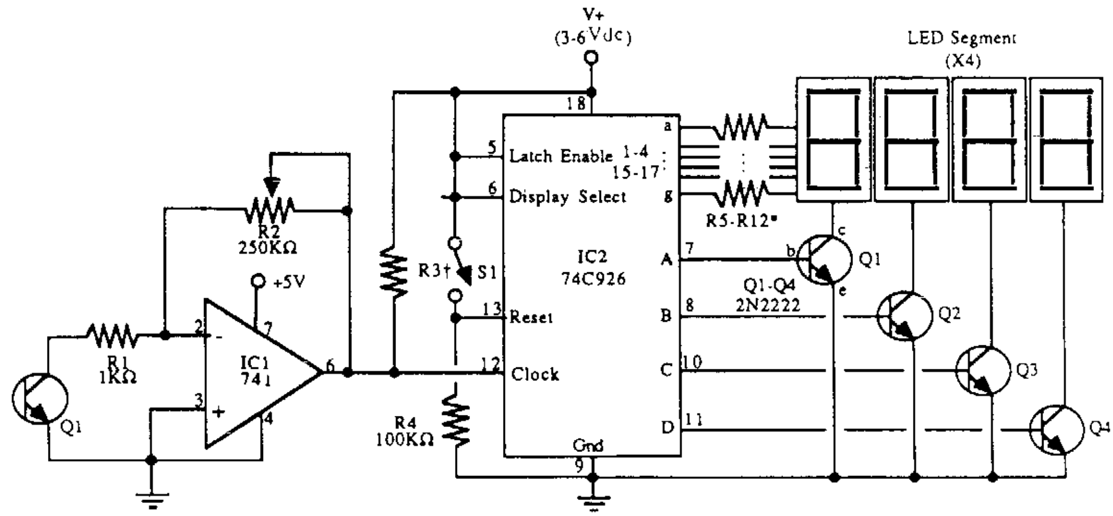
POPULAR ELECTRONICS *Fig. 47-3*

The sun tracker uses a combination of three photoresistors. R7, R8, and R9, to ensure that the circuit will follow the sun during the day, but not look for it at night. Photoresistor cells, R7, R8, and R9 have a value of 160 Ω in full sunlight and 4880 Ω in the shade, that is not absolutely critical. R7 is mounted in a "well" with a narrow slit so that sunlight falls upon it only when the photoresistor is pointed directly at the sun. When that occurs, the resistance of R7 drops. That photoresistor and potentiometer R4 form a voltage divider at the base of the Darlington transistor, Q1. When R7's resistance is low, Q1 will be kept off.

When the sun swings a little westward, R7 will no longer be in sunlight, causing its resistance to go up, which raises the base voltage of Q1 and turns that Darlington on. That, in turn closes the relay, K1, providing current to the drive motor, MOT1, which is a 1.5-Vdc, low-torque hobby motor. The motor then turns slowly (resistor R6 limits the maximum current to the motor and keeps it from running too fast), putting R7 in direct sunlight again; Q1's base voltage then drops and the tracker stops. That is repeated again and again as the sun moves across the sky. Photoresistor R8 is mounted on the outside of the well so that it receives a wide angle of full sunlight. When the sun is shining, R8's resistance is low, keeping Q2 turned off, and allowing the tracker to act as described, without interference. But if the sun "slips" behind a cloud, R8's resistance goes high, producing a forward bias on the base of Q2. That turns that transistor on and sinks the base of Q1 to near ground so that Q1 then remains off. That immobilizes the tracker drive; that also keeps the drive shut down in the dark of night.

Photoresistor, R9, is the dawn sensor. It is mounted on the back of the sun tracker. When the tracker stops at sunset, pointing toward the west, R9 is pointing toward the east. When the sun rises the following morning and shines on R9, its resistance goes low, turning Q2 off and allowing Q1's base to go high. That presents current to the relay and therefore to the drive motor, causing the tracker to swing around to the east.

OPTICAL FRINGE COUNTER

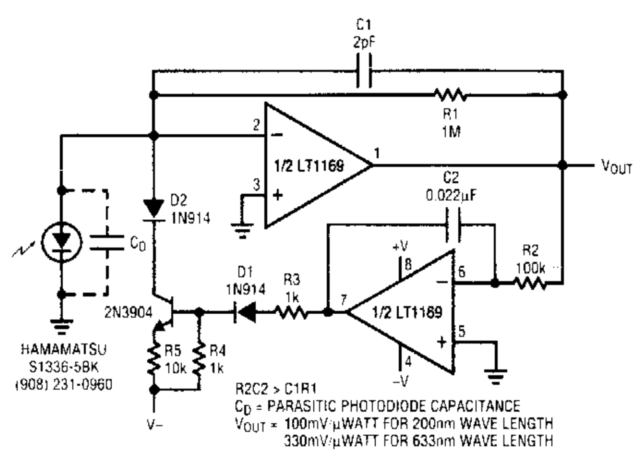


* Not required if +V is 4V or less
† Optional, 10KΩ to 10MΩ, for sensitivity

McGRAW-HILL *Fig. 47-4*

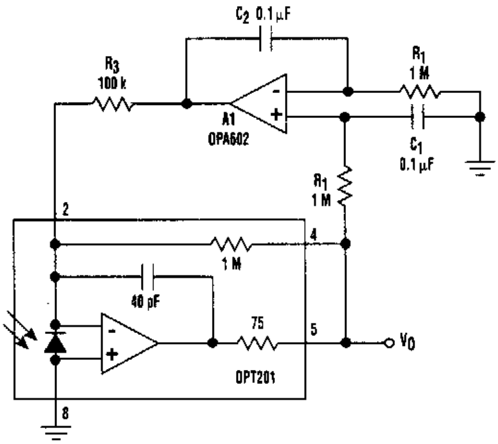
For work with interferometer and optical experiments, this fringe counter can be useful. Photo transistor Q1 provides light and dark sensing. As the sensor is moved across the fringe pattern alternate light and dark areas translate to an electrical waveform. This is amplified by IC1 and counted by IC2. A Schmitt trigger circuit can be added, if desired.

LOW-NOISE LIGHT SENSOR WITH dc SERVO



LINEAR TECHNOLOGY *Fig. 47-5*

PHOTODIODE AMPLIFIER

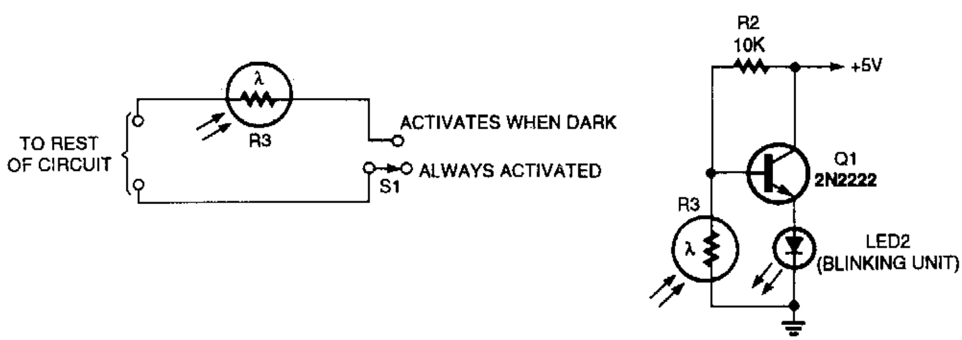


ELECTRONIC DESIGN

Fig. 47-6

A photodiode amplifier combined with a dc-restoration circuit will reject low-frequency ambient background light, easing measurement of a light signal.

LIGHT-SWITCHED LED BLINKER

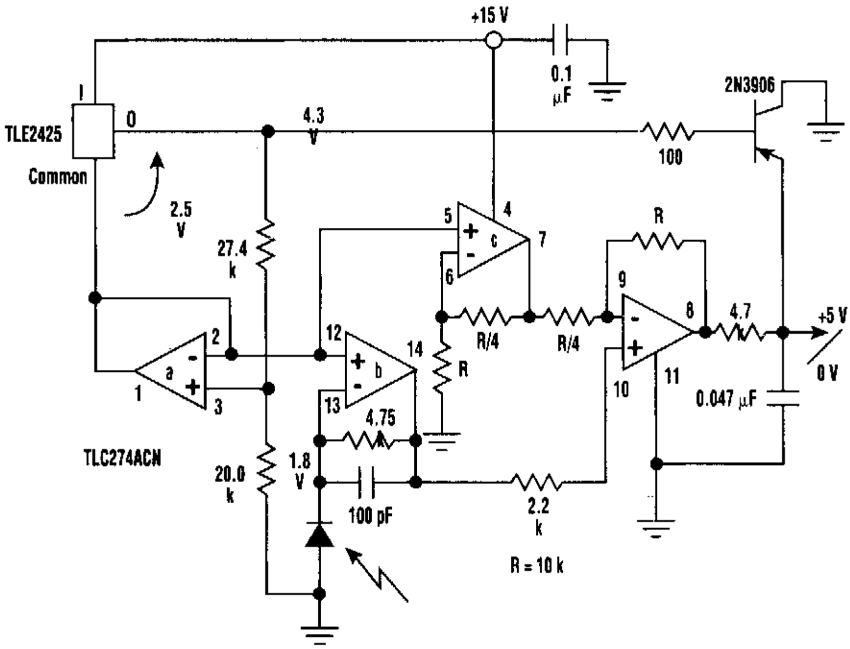


POPULAR ELECTRONICS

Fig. 47-7

This circuit can be used to flash an LED during periods of darkness. Use it for burglar alarm simulators for boats, docks, autos, etc.

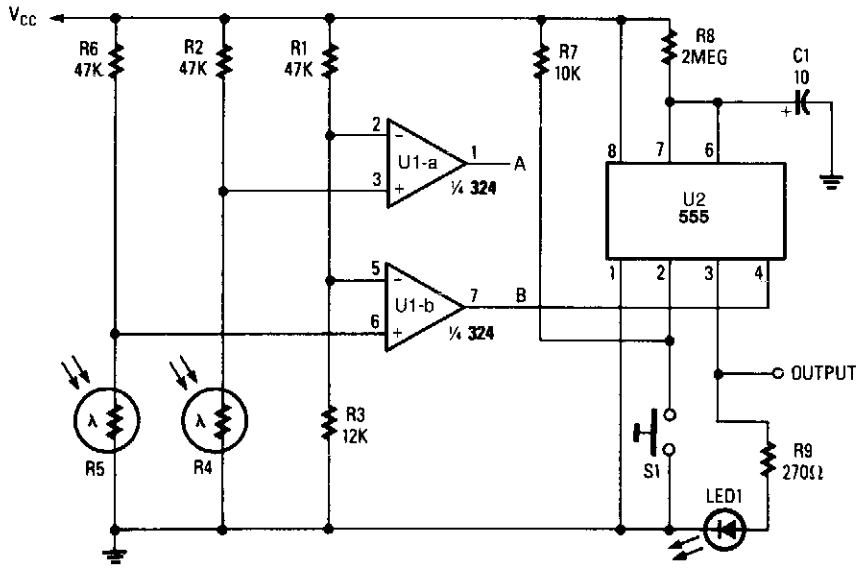
SINGLE-SUPPLY PHOTODIODE AMPLIFIER



ELECTRONIC DESIGN *Fig. 47-8*

This circuit provides a reverse-bias operating point and output voltage offset and uses a single-polarity power supply. The floating reference voltage from TLE2425 serves to bias the diode in a reverse-polarity mode. It also provides a clamping level at the output. Consequently, linear response to illumination is maintained for a 5-V range from dark current to full sunlight conditions.

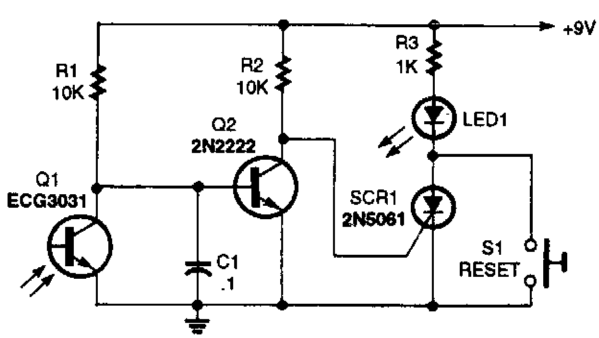
LIGHT-CONTROLLED MONOSTABLE



POPULAR ELECTRONICS *Fig. 47-9*

The light-controlled monostable was produced by combining a 555 monostable multivibrator with a pair of light-controlled comparators. The circuit can be used to enable the operation of the load device, depending on the time of day. During the daylight hours, the timer U2, is disabled, and so produces no output. However, during the nighttime hours, U2 is enabled by the output of U1-b so that pressing S1 initiates a timing cycle, which activates LED1 for a time determined by R8 and C1.

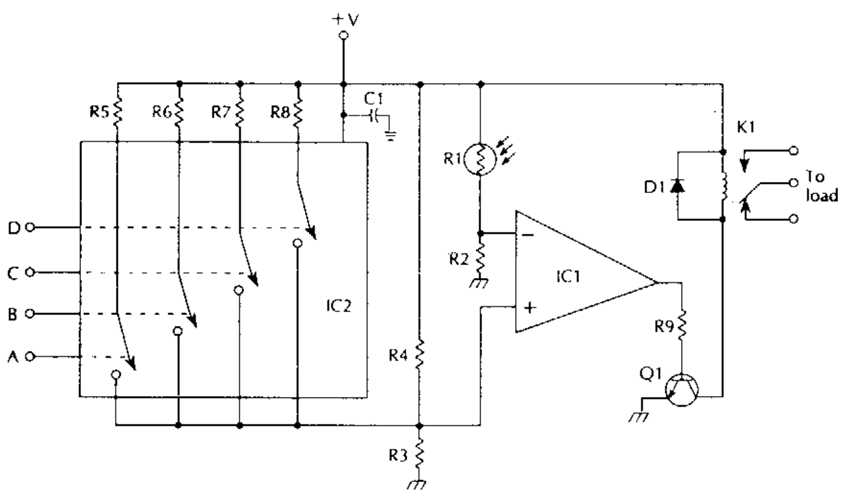
DARKNESS MONITOR



POPULAR ELECTRONICS *Fig. 47-10*

When light strikes detector Q1, Q2 is cut off, allowing bias to reach SCR1, triggering SCR1 and lighting LED1. S1 resets the circuit.

PROGRAMMABLE LIGHT-ACTIVATED RELAY



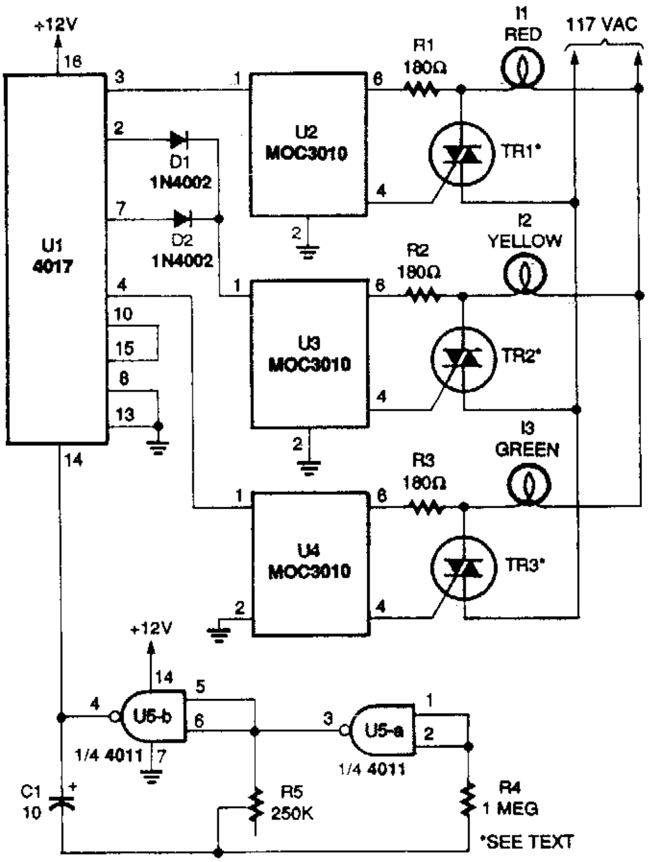
McGRAW-HILL

Fig. 47-11

Digital inputs A, B, C, D select different light levels by varying the value of bridge circuit resistance R_4.

IC1	741 op amp
IC2	CD4066 quad bilateral switch
Q1	NPN transistor (2N2222, 2N3904, or similar)
D1	diode (1N4002, or similar)
C1	0.1-μF capacitor
R1	photoresistor
R2, R3	390-kΩ, ¼-W 5% resistor
R4, R5	1-MΩ, ¼-W 5% resistor
R6	820-kΩ, ¼-W 5% resistor
R7	470-kΩ, ¼-W 5% resistor
R8	270-kΩ, ¼-W 5% resistor
R9	100-kΩ, ¼-W 5% resistor
K1	relay to suit load

TRAFFIC LIGHT CONTROLLER



POPULAR ELECTRONICS *Fig. 47-12*

Oscillator U5A-B drives a 4017 divide-by-10 counter. The first output of U1 appears at pin 3, which supplies a positive voltage to U2, a MOC 3010 optocoupler/triac-driver IC, turning it and triac TR1 on. That lights I1, the red lamp. The second output appears at pin 2 and passes through D1 to the second MOC 3010, U3, thereby lighting the yellow lamp, I2. The third output at pin 4 turns on UJ4 and the green lamp, I3. The fourth output at pin 7 travels through D2 and into U3 to light the yellow lamp, I2, again.

If you would like the traffic-light system to follow the normal sequence of green, yellow, and red, make the following circuit changes: Disconnect pins 10 and 15 of U1 from each other. Remove D1 and D2 and connect pin 2 of U1 to pin 1 of U3. Then connect pins 7 and pins 15 of U1 together. Use U2 to drive I3 (the green light) and U4 for I1 (the red light).

COLORIMETER

TABLE 1—LED COLORS AND CODES

LED	Wave-length	Color	Activation Value
LED1	470 nm	Blue	2^0=1
LED2	560 nm	Green	2^1=2
LED3-a	590 nm	Yellow	2^2=4
LED3-b	700 nm	Red	2^3=8
LED4	630 nm	Orange	2^4=16
LED5	665 nm	Crimson	2^5=32
LED6	482 nm	Aqua	2^6=64

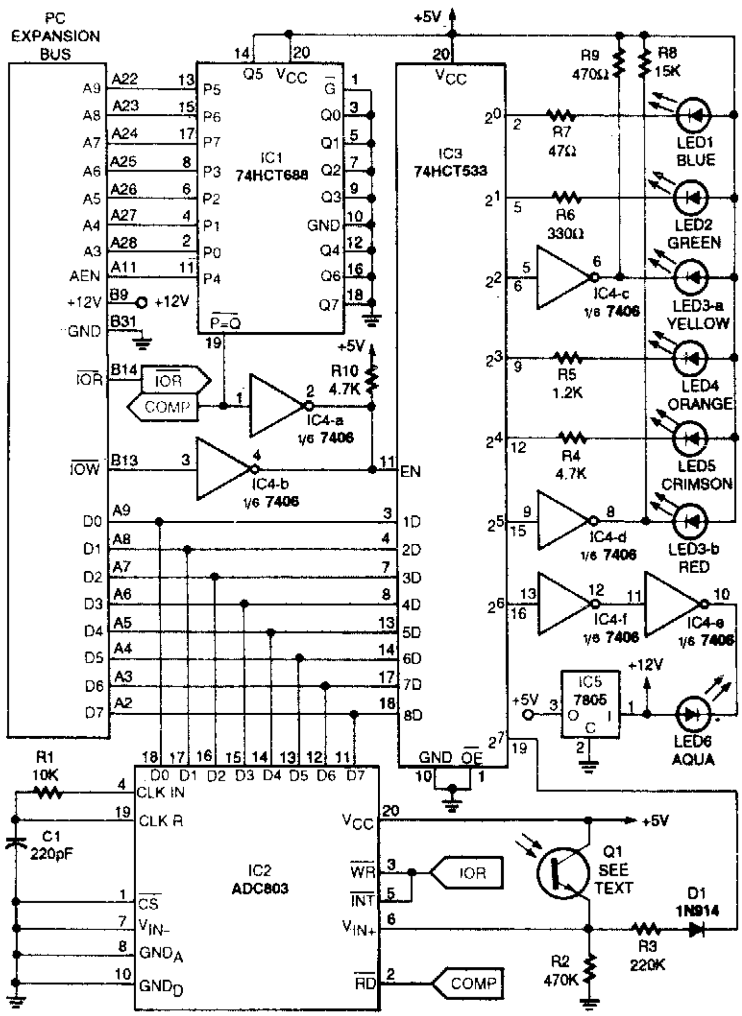


Fig. 47-13

LISTING 1—CALIBRATION PROGRAM

```
10 'CALIBRAT.BAS calibration program
20 CLS:KEY OFF:N=0:ADR=512:OPEN"R",1,"CAL1",16:OPEN"r",2,"cal2",24
30 FIELD 1,2AS B$,2AS G$,2AS Y$,2AS O$,2AS C$,2AS R$,2AS A$,2AS AG$
40 FIELD 2,24AS ID$
50 PRINT "reference number",N+1:OUT ADR,255:BEEP:INPUT "Enter Name of Standard
or 'E' To End";TEMPID$
60 IF TEMPID$="E" OR TEMPID$="e" THEN N=0: GOTO 200
70 IF TEMPID$="n" THEN INPUT"enter n to redo ",N:N=N-1:GOTO 50
80 N=N+1:FOR H=0 TO 7:K=0:IF H<7 THEN Z=2^H ELSE Z=194
90 OUT ADR,Z:FOR I=1 TO 500:NEXT I
100 FOR J=1 TO 50:K=K+INP(ADR):NEXT J
110 IF H=0 THEN LSET B$=MKI$(K)
120 IF H=1 THEN LSET G$=MKI$(K)
130 IF H=2 THEN LSET Y$=MKI$(K)
140 IF H=3 THEN LSET O$=MKI$(K)
150 IF H=4 THEN LSET C$=MKI$(K)
160 IF H=5 THEN LSET R$=MKI$(K)
170 IF H=6 THEN LSET A$=MKI$(K)
180 IF H=7 THEN LSET AG$=MKI$(K)
190 NEXT H:LSET ID$=TEMPID$:PUT 1,N:PUT 2,N:CLS:GOTO 50
200 N=N+1:GET #1,N:GET #2,N:IF N>(LOF(1)/16) THEN END
210 B=CVI(B$):G=CVI(G$):Y=CVI(Y$):O=CVI(O$):C=CVI(C$):R=CVI(R$):A=CVI(A$):
AG=CVI(AG$)
220 PRINT N,ID$:GOTO 200
```

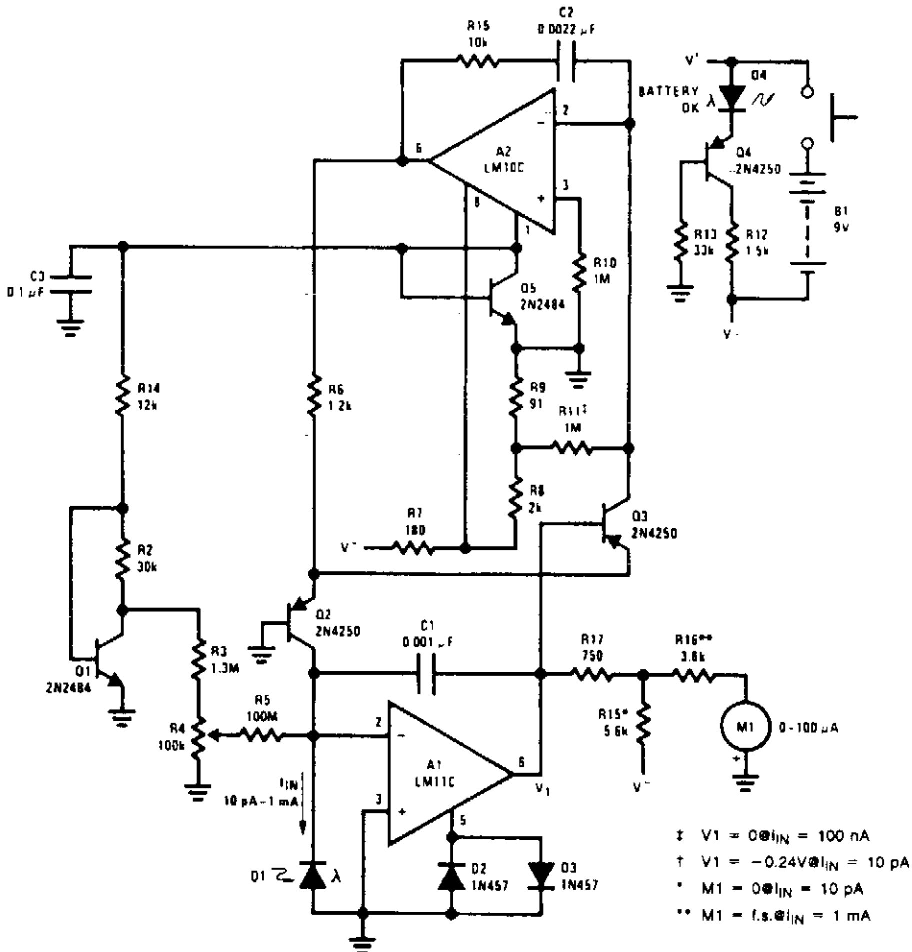
LISTING 2—IDENTIFICATION PROGRAM

```
1 'IDENTIFY.BAS  identification program
10 ADR=512:OUT ADR,255:PRINT:INPUT "Hit Enter To Scan/Identify Unknown
Color";A
20 IF A=9 THEN RUN"fcal"
30 ERP=1E+20:OPEN"R",1,"call",16
40 FOR H=O TO 7:K=0:IF H<7 THEN Z=2^H ELSE Z=194
50 OUT ADR,Z: FOR I=1 TO 500:NEXT I
60 FOR J=1 TO 50:K=K+INP(ADR):NEXT J
70 IF H=0 THEN BU=K ELSE IF H=1 THEN GU=K ELSE IF H=2 THEN YU=K
80 IF H=3 THEN OU=K ELSE IF H=4 THEN CU=K ELSE IF H=5 THEN RU=K
90 IF H=6 THEN AU=K ELSE IF H=7 THEN AGU=K
100 NEXT H:BEEP
110 OUT 512,255:OPEN"r",2,"cal2",24:FIELD 1,2AS B$,2AS G$,2AS Y$,2AS O$,2AS
C$,2     AS R$,2AS A$,2AS AG$:B=LOF(1)/16
120 FOR N=1 TO B:GET #1,N:IF ABS( CVI(B$)-BU)>400 THEN 140
130 ER=(CVI(B$)-BU)^2+(CVI(G$)-GU)^2+(CVI(Y$)-YU)^2+(CVI(O$)-OU)^2+(CVI(C$)-
CU)^     2+(CVI(R$)-RU)^2+1*((CVI(A$)-AU)^2)+2*((CVI(AG$)-AGU)^2):IF ER<ERP
THEN ERP=     ER:NN=N
140 NEXT N
150 FIELD 2, 24AS ID$: GET #2,NN
160 CLS:PRINT "Best Color Match",ID$:PRINT"Relative Error",ERP:PRINT"reference
number",NN:RUN
```

A hardware/software combination activates. In turn, one of several LEDs emits a portion of the visible spectrum. A phototransistor measures the light reflected by the surface being measured, and an 8-bit analog-to-digital converter (ADC) translates the phototransistor's output into a digital format that the computer can interpret. Seven LEDs (blue, aqua, green, yellow, orange, crimson, and red) provide a range of readings across the visible spectrum. Lack of spectral continuity among adjacent LED colors could skew results, so the circuit provides built-in compensation for this error.

Two simple BASIC programs control the circuit's operation. One allows you to define a set of standards by measuring known color samples and recording the values with an associated name. The other program measures unknown samples and provides the best match with the defined standards, as well as a relative error factor.

EIGHT-DECADE LIGHT METER

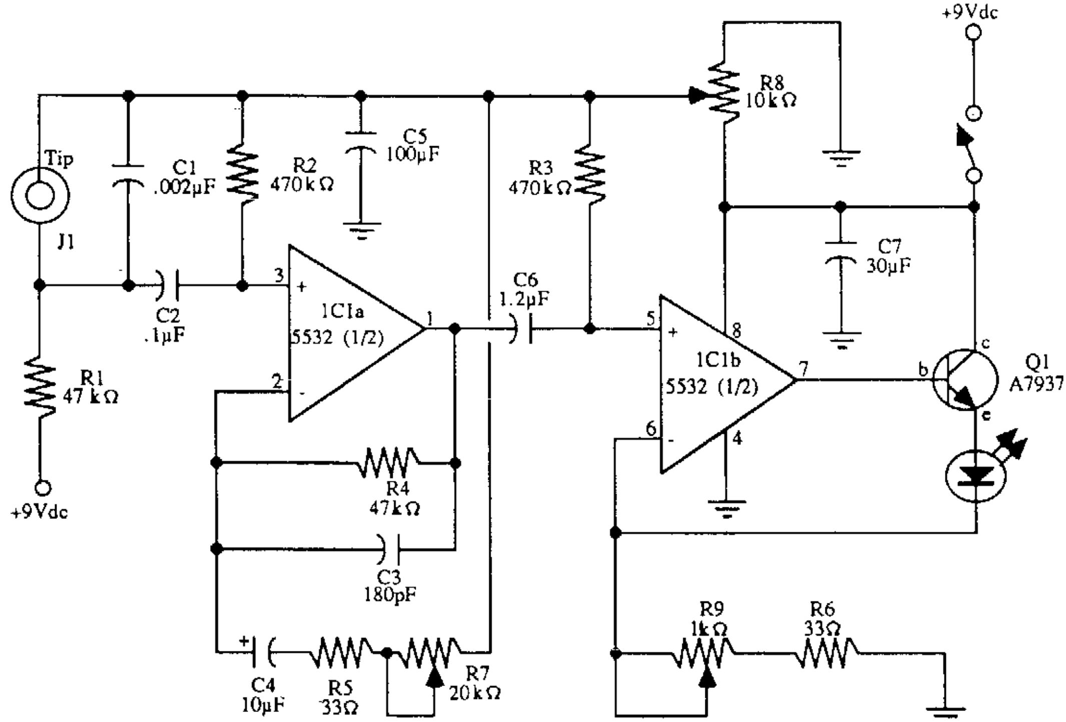


NATIONAL SEMICONDUCTOR *Fig. 47-14*

A logarithmic amplifier is adapted to a battery-powered light meter. An LM10, combined op amp and reference, is used for the second amplifier and to provide the regulated voltage for offsetting the logging circuit and powering the bias-current compensation. This can provide input current resolution of better than ±2 pA over 15 to 55° C. Because a meter is the output indicator, there is no need to optimize frequency compensation. Low-cost single transistors are used for logging because the temperature range is limited. The meter is protected from overloads by clamp diodes D2 and D3.

Silicon photodiodes are more sensitive to infrared than visible light, so an appropriate filter must be used for photography. Alternately, gallium-arsenide-phosphide diodes with suppressed IR response are becoming available.

LED LIGHTWAVE COMMUNICATIONS TRANSMITTER



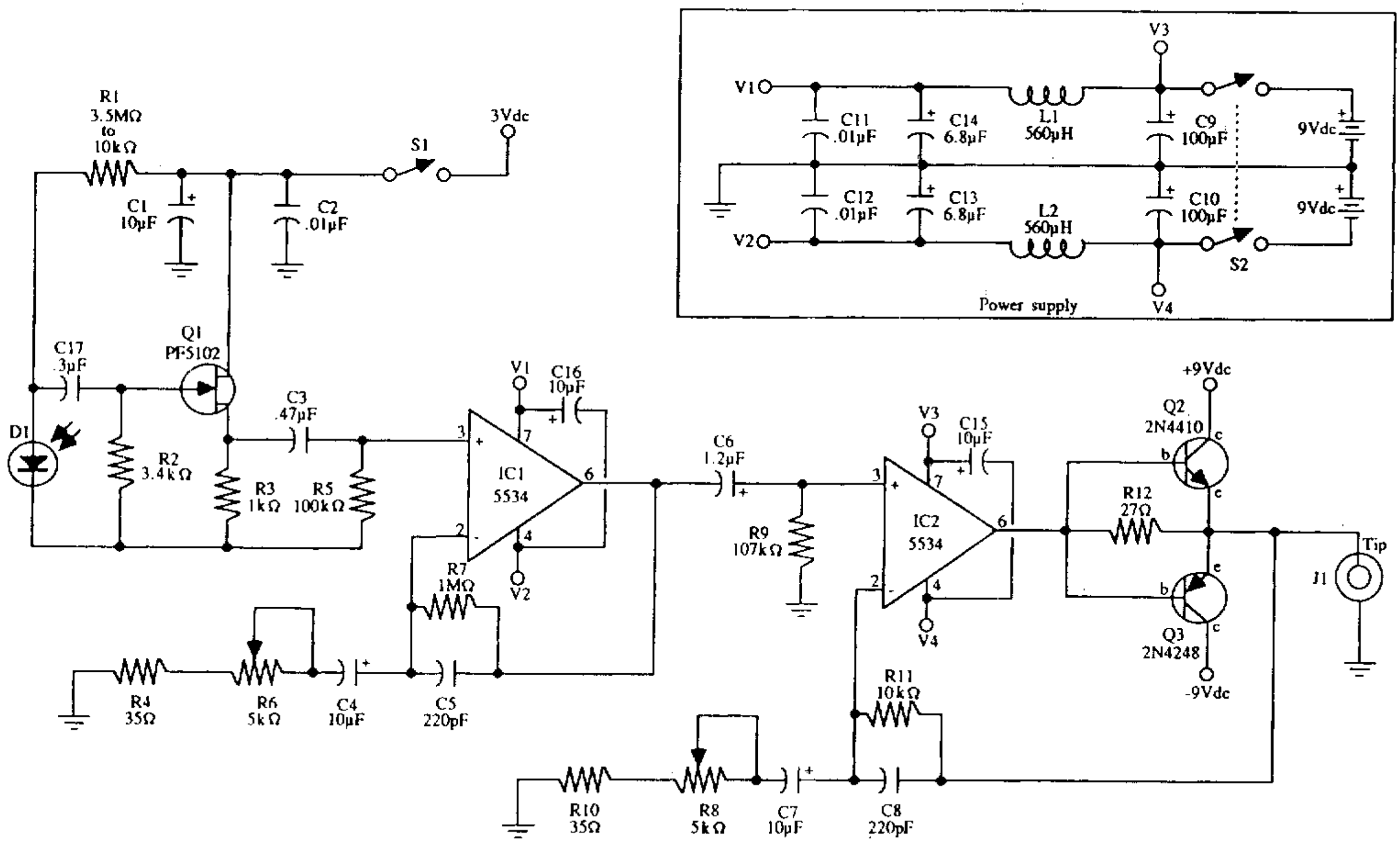
R1,R4	47 kΩ
R2,R3	470 kΩ
R5,R6	33 Ω
R7,R9	1 kΩ potentiometer
R8	10 kΩ potentiometer
C1	0.002 μF disc
C2	0.1 μF disc
C3	180 pF disc
C4	10 μF polarized electrolytic
C5	100 μF polarized electrolytic
C6	1.2 μF polarized electrolytic
C7	30 μF polarized electrolytic
IC1	5532 low-noise amplifier IC
Q1	A7937 transistor
LED1	High-output LED (see text)
J1	Miniature phone jack (for electret condenser microphone)
S1	SPST switch

All resistors are 5 to 10 percent tolerance, 1/4 watt. All capacitors are 10 to 20 percent tolerance, rated at 35 volts or more.

McGRAW-HILL

Fig. 47-15

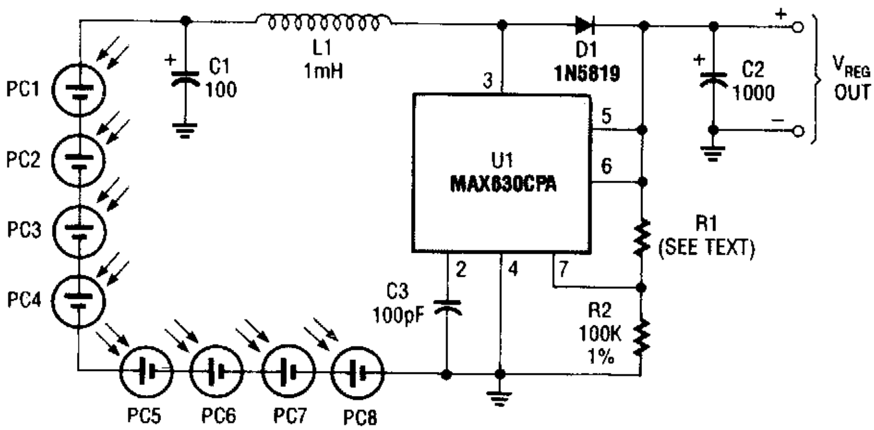
LED LIGHTWAVE RECEIVER



McGRAW-HILL

Fig. 47-16

SOLAR POWER SUPPLY

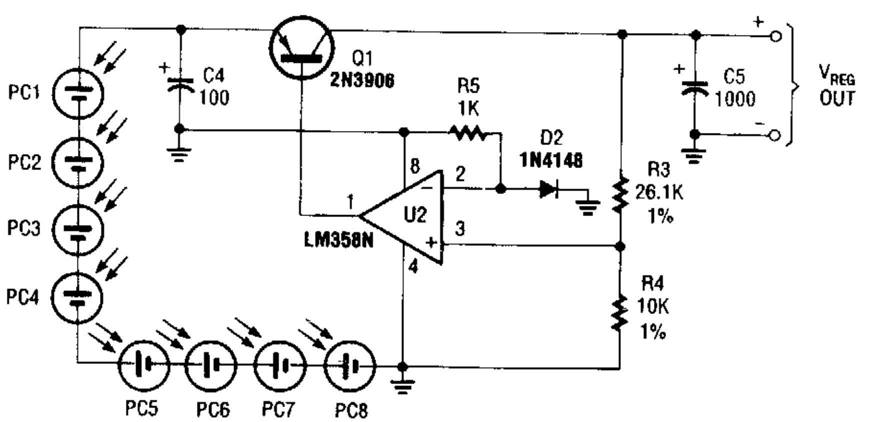


ELECTRONICS HOBBYIST HANDBOOK

Fig. 47-17

This circuit delivers either 4.8 or 7.2 V regulated at 15 mA with a 3-V input from a bank of photocells. R1 should be 453 kΩ for a 7.2-V output and 274 kΩ for a 4.8-Vdc output. Regulator efficiency is around 70%. This should be considered when selecting suitable solar cells.

SOLAR POWER SUPPLY WITH LINEAR REGULATOR

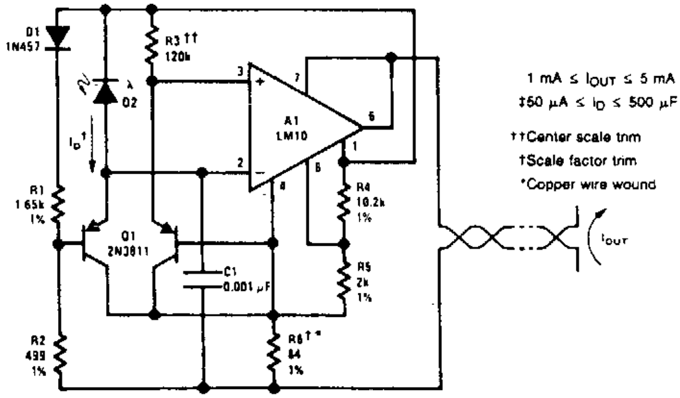


ELECTRONICS HOBBYIST HANDBOOK

Fig. 47-18

This regulator delivers a constant 2.4 Vdc for powering small devices that run on two AA cells, such as cassettes and small radios. Regulator drop is about 0.3 volt. This should be considered when choosing solar cells. Load current is typically 125 mA.

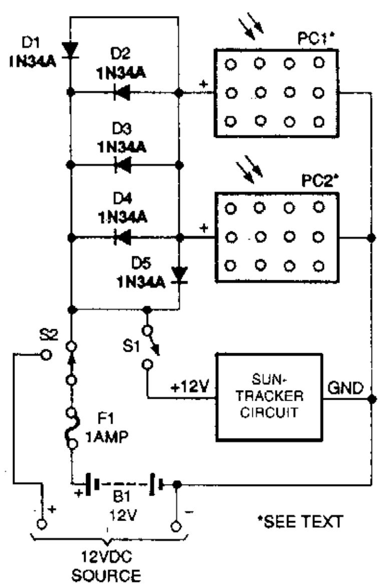
PHOTODIODE LOG CONVERTER/TRANSMITTER



NATIONAL SEMICONDUCTOR ***Fig. 47-19***

A logarithmic conversion is made on the output current of a photodiode to compress a four-decade, light-intensity variation into a standard transmission range. The circuit is balanced at mid-range, where R3 should be chosen so that the current through it equals the photodiode current. The log-conversion slope is temperature compensated with R6. Setting the reference output to 1.22 V gives a current through R2 that is proportional to absolute temperature because of D1 so that this level-shift voltage matches the temperature coefficient of R6. C1 has been added so that large-area photodiodes with high capacitance do not cause frequency instabilities.

RECHARGEABLE SOLAR POWER FOR SUN TRACKER



POPULAR ELECTRONICS ***Fig. 47-20***

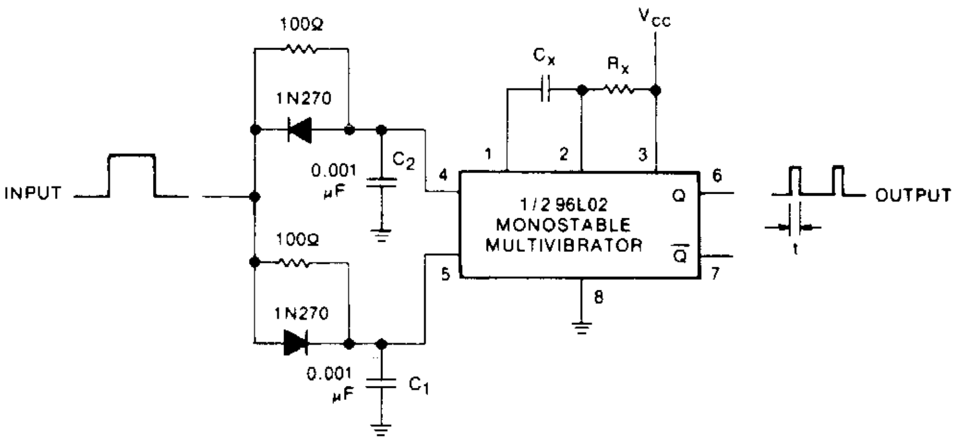
This application circuit provides rechargeable solar power for the sun tracker, as well as for another 12-volt device. PC1 and PC2 can be mounted on sun tracker assembly.

48

Logic Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOGIC STATE-CHANGE INDICATOR



NASA TECH BRIEFS ***Fig. 48-1***

A circuit consisting of a one-shot multivibrator IC, a pair of diodes, and some resistors and capacitors delivers an output pulse when the logic state at its input terminal changes—either from high to low or from low to high. Thus, this circuit can serve as a state-change indicator or as a frequency doubler for a square-wave input.

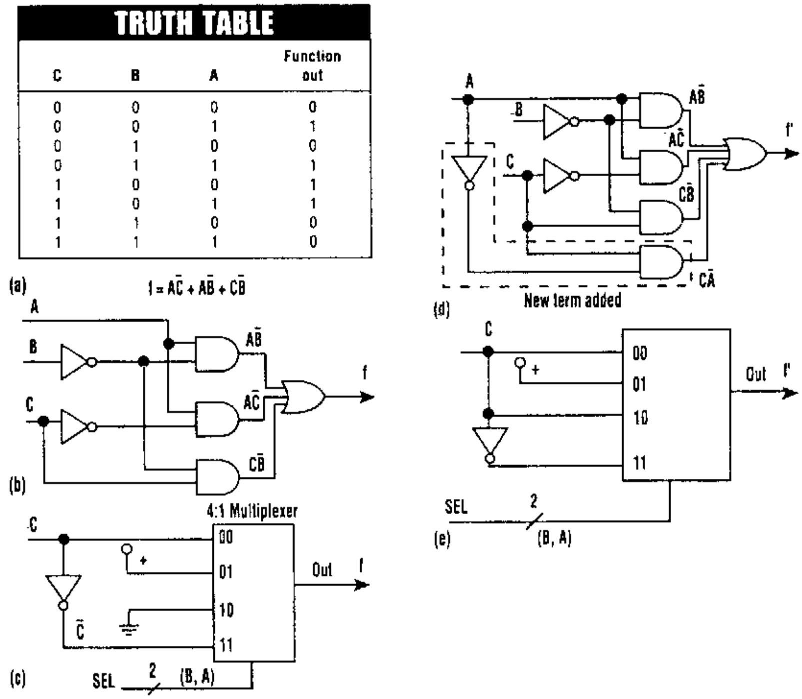
Any monostable can be used; the arrangement in the figure achieves low power dissipation (80 milliwatts) by using half of a Fairchild 96L02 transistor-transistor-logic dual multivibrator. The 96L02 is triggered when pins 3 and 5 are high and pin 4 changes state from low to high. It also triggers if pin 3 is high, pin 4 is low.

The circuit shown here allows these conditions to be satisfied with a single input terminal, plus the fixed bias on pin 3; the arrangement of resistors, capacitors, and diodes automatically biases pin 5 high when transmitting a rising transition to pin 4 and biases pin 4 low when applying a falling transition to pin 5.

For example, if the input terminal has been low and then goes high, C1 charges through a forward-biased diode that shunts its 100-Ω resistor; therefore pin 5 goes high immediately. C2 charges through 100 Ω; however, because its diode is back-biased, the rising level is not applied to pin 4 until 5 is already high. Therefore, the conditions for triggering an output pulse are satisfied.

The output pulse duration, t, is set by the value of time constant R_xC_x.

COMBINATORIAL LOGIC MULTIPLEXER

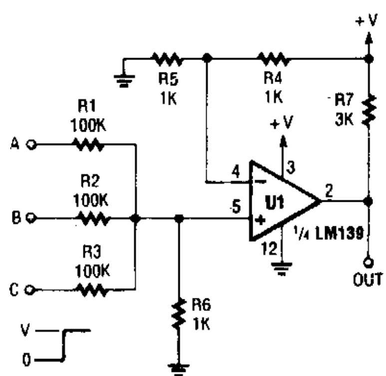


C	B	A	Function out
0	0	0	0
0	0	1	1
0	1	0	0
0	1	1	1
1	0	0	1
1	0	1	1
1	1	0	0
1	1	1	0

ELECTRONIC DESIGN *Fig. 48-2*

Combinatorial logic can be implemented simply by using a multiplexer instead of logic gates. Shown are the truth table (A), its logic circuit (B), and the multiplexer connections (C). If the logic circuitry is changed (D), the multiplexer would be reconnected (E).

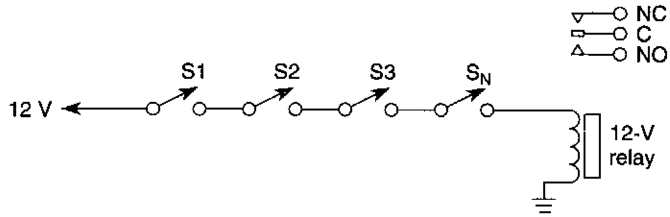
AND GATE



POPULAR ELECTRONICS *Fig. 48-3*

An LM139 is configured as an AND gate (TTL or CMOS is usually used). With this idea, you can use leftover IC sections and save an extra package in some instances.

RELAY "AND" CIRCUIT

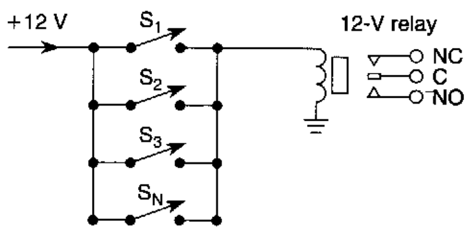


WILLIAM SHEETS

Fig. 48-4

All switches S1 through S_n must be closed to operate the relay. If one opens, the relay drops out. Use this circuit for burglar alarms, etc.

RELAY "OR" CIRCUIT



WILLIAM SHEETS

Fig. 48-5

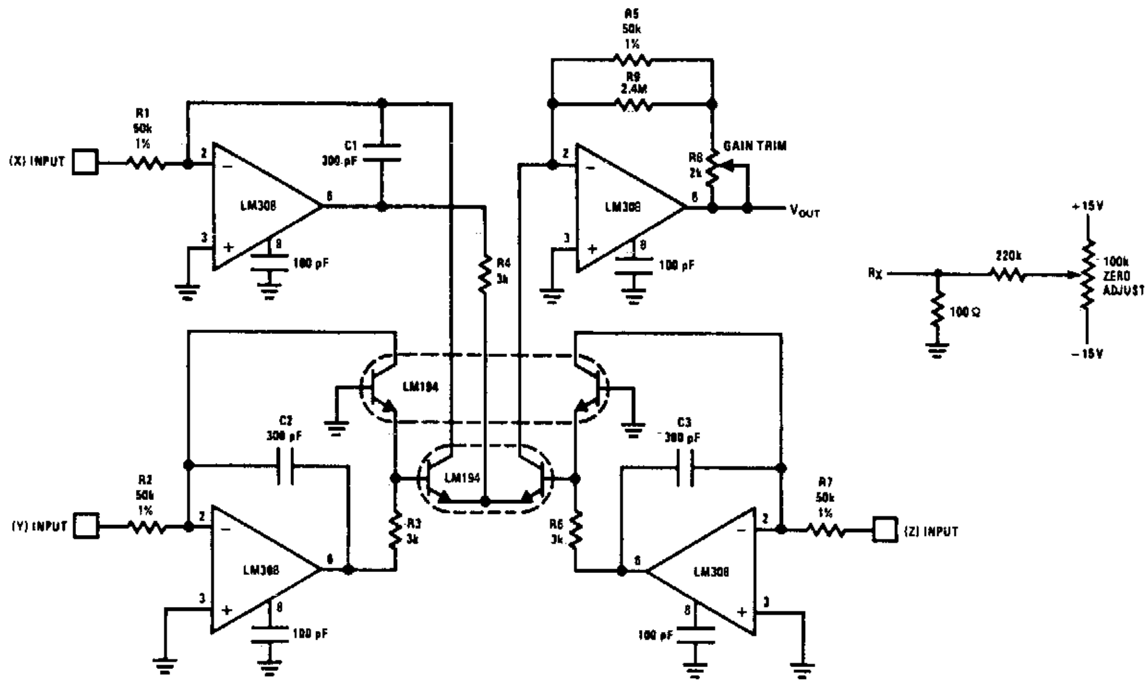
Closing any switch S1, S2, S3, or S_n will actuate the relay (N = any number). Use this circuit for burglar alarms, etc.

49

Mathematical Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOW-COST ONE-QUADRANT MULTIPLIER/DIVIDER



NATIONAL SEMICONDUCTOR

Fig. 49-1

This circuit will produce an output that is proportional to the product of the (*X*) and (*Y*) inputs divided by the *Z* input. All inputs must be positive, limiting operation to one quadrant. For very low level inputs, the offset voltage in the LM308s might create large percentage errors referred to input. A simple scheme for offsetting any of the LM308s to zero is shown in dotted line; the positive input of the appropriate LM308 is simply tied to R_x instead of ground for zeroing. The summing mode of operation on all inputs allows easy scaling on any or all inputs. Simply set the input resistor equal to $(V_{in(max)}/(200\ \mu A)$. V_{out} is equal to:

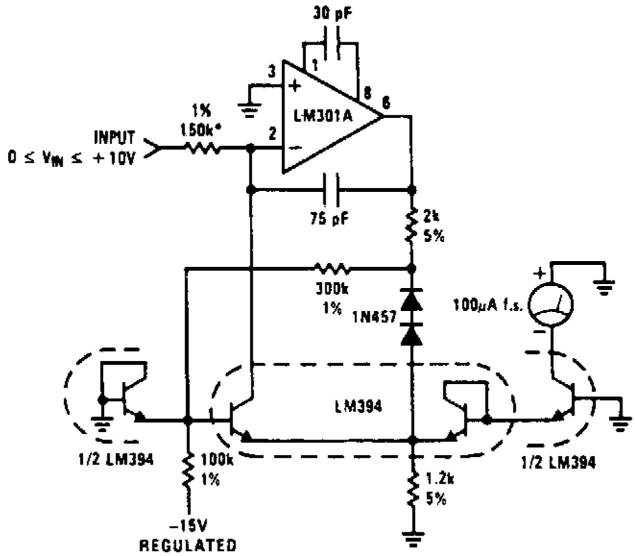
$$V_{out} = \frac{\left(\frac{X}{R_1}\right)\left(\frac{Y}{R_2}\right)(R_5)}{\frac{Z}{R_7}}$$

Input voltages above the supply voltage are allowed because of the summing mode of operation. Several inputs can be summed at *X*, *Y*, and *Z*.

For a simple (*X*) • (*Y*) or (*X*)/*Z* function, the unused input must be tied to the reference voltage. Perturbations in this reference will be seen at the output as scale factor changes, so a stable reference is necessary for precision work. For less critical applications, the unused input can be tied to the positive supply, with

$$R = \frac{V+}{200\ \mu A}$$

LOW-COST ACCURATE SQUARE-ROOT CIRCUIT



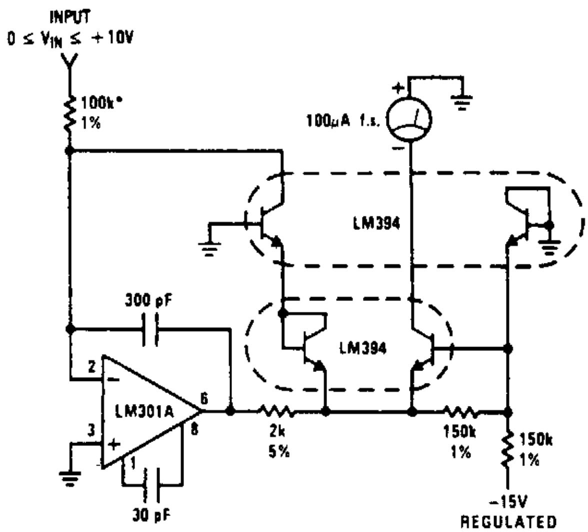
*Trim for full scale accuracy.

NATIONAL SEMICONDUCTOR

Fig. 49-2

The circuit will generate a square-root function, accurately and inexpensively. The output is a current that can be used to drive a meter directly or be converted to a voltage with a summing junction current-to-voltage converter. The −15-V supply is used as a reference, so it must be stable. A 1% change in the −15-V supply will give a ½% shift in output reading. No positive supply is required when an LM301A is used because its inputs can be used at the same voltage as the positive supply (ground). The two 1N457 diodes and the 300-kΩ resistor are used to temperature-compensate the current through the diode-connected ½LM394.

LOW-COST ACCURATE SQUARING CIRCUIT



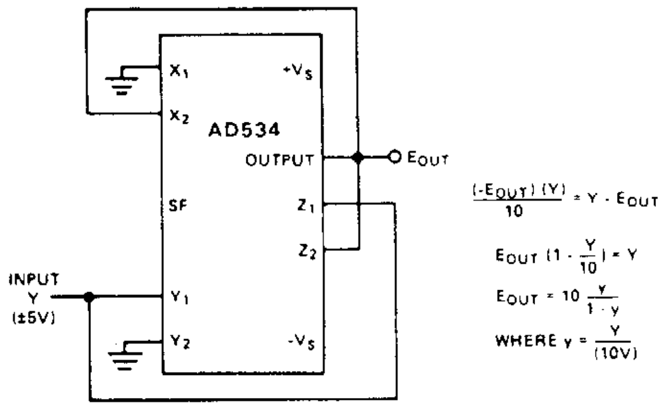
*Trim for full scale accuracy.

NATIONAL SEMICONDUCTOR

Fig. 49-3

The circuit shown will square the input signal and deliver the result as an output current. Full-scale input is 10 V, but this can be changed simply by changing the value of the 100-kΩ input resistor. As in the square root circuit, the –15-V supply is used as the reference. In this case, however, a 1% shift in supply voltage produces a 1% shift in the output signal. The 150-kΩ resistor across the base-emitter of ½LM394 provides slight temperature compensation of the reference current from the –15-V supply. For improved accuracy at low input signal levels, the offset voltage of the LM301A should be zeroed out, and a 100-kΩ resistor should be inserted in the positive input to provide input to provide optimum dc balance.

BRIDGE LINEARIZING FUNCTION



ANALOG DEVICES

Fig. 49-4

If one arm of a Wheatstone Bridge varies from its nominal value by a factor, $(1 + 2x)$, the voltage or current output of the bridge will be (with appropriate polarities):

$$y \approx \frac{x}{1+x}$$

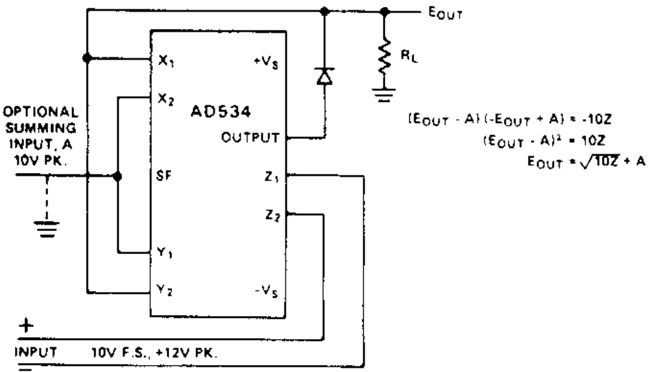
Linear response requires very small x and, usually, preamplification. The circuit shown here enables large-deviation bridges to be used without losing linearity.

The circuit computes the inverse of the bridge function, i.e.,

$$x \approx \frac{y}{1+y}$$

Depending on which arm of the bridge varies, it might be necessary to reverse the polarity of the z connections.

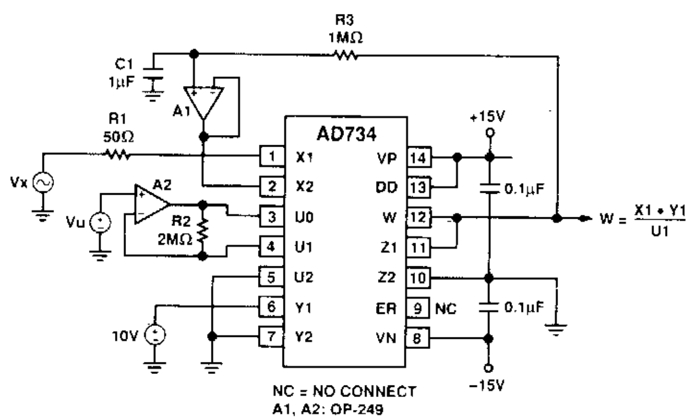
SQUARE ROOTER



ANALOG DEVICES **Fig. 49-5**

This illustration shows the connection of the AD534 for square-rooting, with differential inputs. The diode prevents a latching condition—common to this configuration—which would occur if the input momentarily changed polarity. As shown, the output is always positive; it can be changed to a negative output by reversing the diode polarity and interchanging the X inputs. Because the signal input is differential, all combinations of input and output polarities can be realized. If the output circuit does not provide a resistive load to ground, one should be connected to maintain diode conduction. For critical applications, the Z offset can be adjusted for greater accuracy below 1 V.

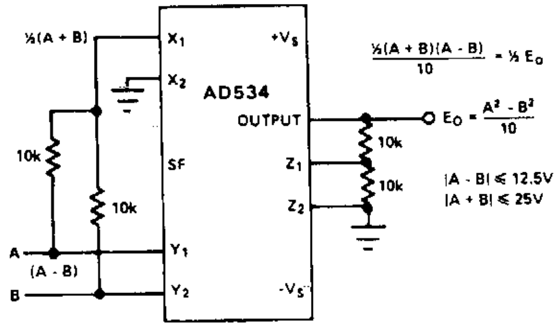
ANALOG VARIABLE MULTIPLIER/DIVIDER



ANALOG DEVICES **Fig. 49-6**

An output voltage ($W = X_1 \times Y_1/U_1$) is produced by this multiplier circuit. The AD734 is a four-quadrant multiplier.

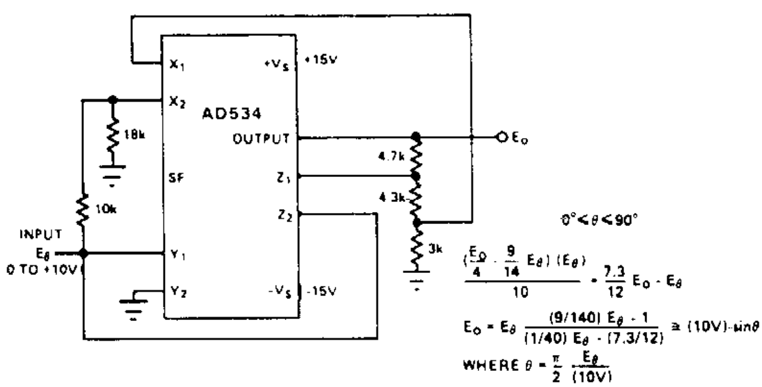
DIFFERENCE OF SQUARES



ANALOG DEVICES *Fig. 49-7*

A single AD534 can be used to compute the difference of the squares of two input signals. The function can be useful in vector computations, and in weighting the difference of two magnitudes to emphasize the greater nonlinearity.

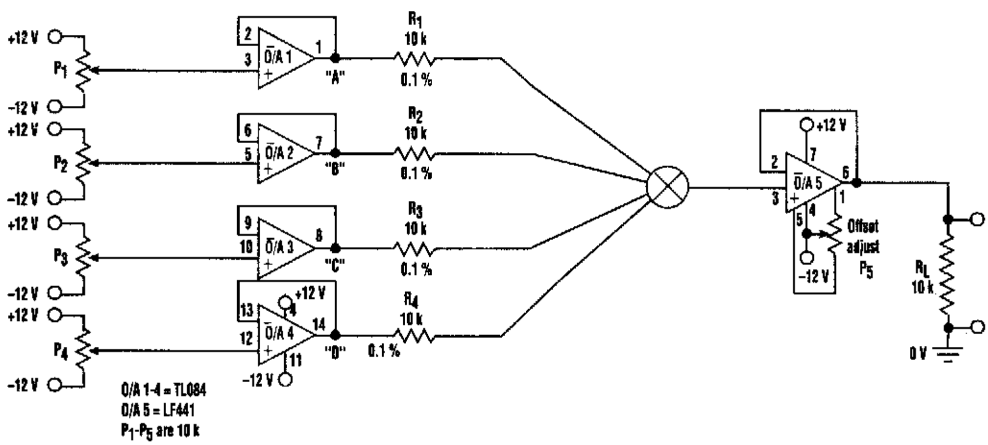
APPROXIMATION FOR SIN ϕ



ANALOG DEVICES *Fig. 49-8*

The AD534 is remarkably easy to use in the implementation of the approximation formulas described in Chapter 2-1 of the *Nonlinear Circuits Handbook*. Many of these involve implicit loops to generate the function and previously required several additional op amps for the addition and subtraction of the various terms. This circuit is an example of what can be done with external resistors only. For ϕ between 0° and 90°, the approximation maintains a theoretical accuracy to within 0.5% of full-scale; 0.75% is practical with AD534L and 0.1% resistances were used.

SIMPLE ANALOG AVERAGING CIRCUIT



	Voltage at points	TABLE							
	"A"	+5	+5	-5	+1.5	-1	+8	+10	+10
	"B"	+5	-5	-5	-3	-3	+8	+10	-9
	"C"	+5	+5	+5	-6	-5	-3	+10	+10
	"D"	+5	+5	+5	+2.5	-8	+9	+9	+10
Output across R_L		+5	+2.5	0	-1.25	-4.25	+5.5	+9.75	+5.25

ELECTRONIC DESIGN

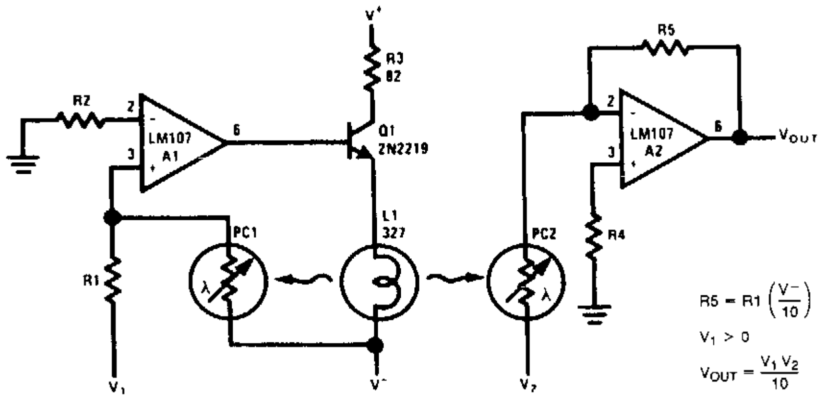
Fig. 49-9

At times, an analog circuit that averages rather than sums can be quite handy. You won't usually find this type of circuit in op-amp books, possibly because the op amp is used only as a buffer. For best accuracy, an FET should be used with the offset adjusted out. In addition, the "averaging" resistors (R1 through R4) should be of close tolerance.

Looking at the test circuit, op amps 1 through 4 are used to alleviate interaction between adjustment potentiometers P1 through P4 and so that R1 through R4 see the same low impedance.

The table shows some arbitrarily set voltages and the resulting output voltage across R_L.

SIMPLE ANALOG MULTIPLIER



NATIONAL SEMICONDUCTOR

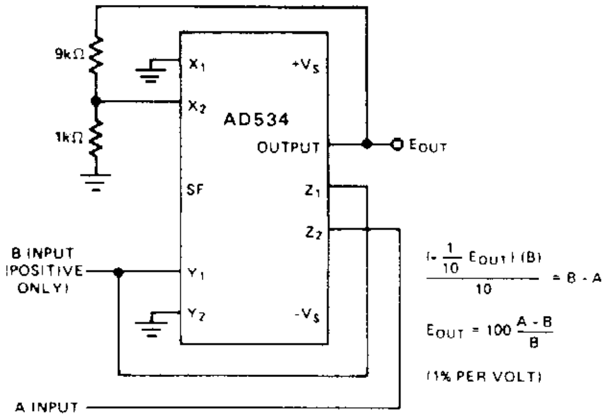
Fig. 49-10

Circuit operation can be understood by considering A2 as a controlled-gain amplifier, amplifying V_2, whose gain depends on the ratio of the resistance of PC2 to R5 and by considering A1 as a control amplifier, which establishes the resistance of PC2 as a function of V1. In this way, V_{OUT} is a function of both V_1 and V_2.

A1, the control amplifier, provides drive for the lamp, L1. When an input voltage, V_1, is present, L1 is driven by A1 until the current to the summing junction from the negative supply through PC1 is equal to the current to the summing junction from V1 through R1. Because the negative supply voltage is fixed, this forces the resistance of PC1 to a value that is proportional to R1 and to the ratio of V_1 to $V-$. L1 also illuminates PC2 and, if the photoconductors are matched, causes PC2 to have a resistance equal to PC1.

A2, the controlled gain amplifier, acts as an inverting amplifier whose gain is equal to the ratio of the resistance of PC2 to R5. If R5 is chosen equal to the product of R_1 and $V-$, then V_{OUT} becomes simply the product of V_1 and V_2. R5 can be scaled in powers of 10 to provide any required output scale factor.

Δ% RATIO COMPUTER



ANALOG DEVICES *Fig. 49-11*

The percentage-deviation function is of practical value for many applications in measurement, testing, and control. For example, the output of this circuit might be applied to a pair of biased comparators to stimulate particular actions or displays, depending on whether the gain of a circuit under test were within limits, or deviating by a preset amount in either direction.

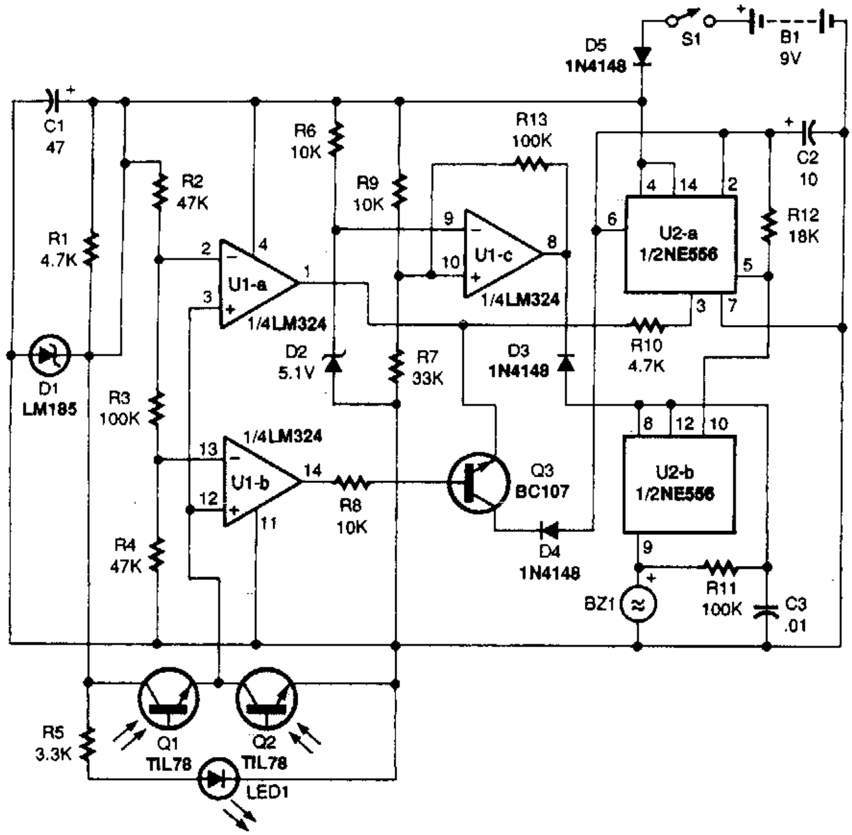
The indicated scale factor, 1%/V, is convenient. However, other sensitivities, from 10%/V to 0.1%/V, as required by the application, can be obtained by altering the feedback attenuation ratio, from 1 to 1/100. Gain or attenuation is easily applied to the A signal externally for calibration to the normalized form.

50

Measuring and Test Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC LEVEL



POPULAR ELECTRONICS ***Fig. 50-1***

The electronic level uses a pair of phototransistors and an infrared LED to sense bubble position. In this circuit, the amounts of infrared radiation received by phototransistors Q1 and Q2 are translated by op-amp U1 and dual-timer U2 into either a steady tone, or a fast- or slow-pulsing one.

SINGLE-CHIP DIGITAL VOLTMETER

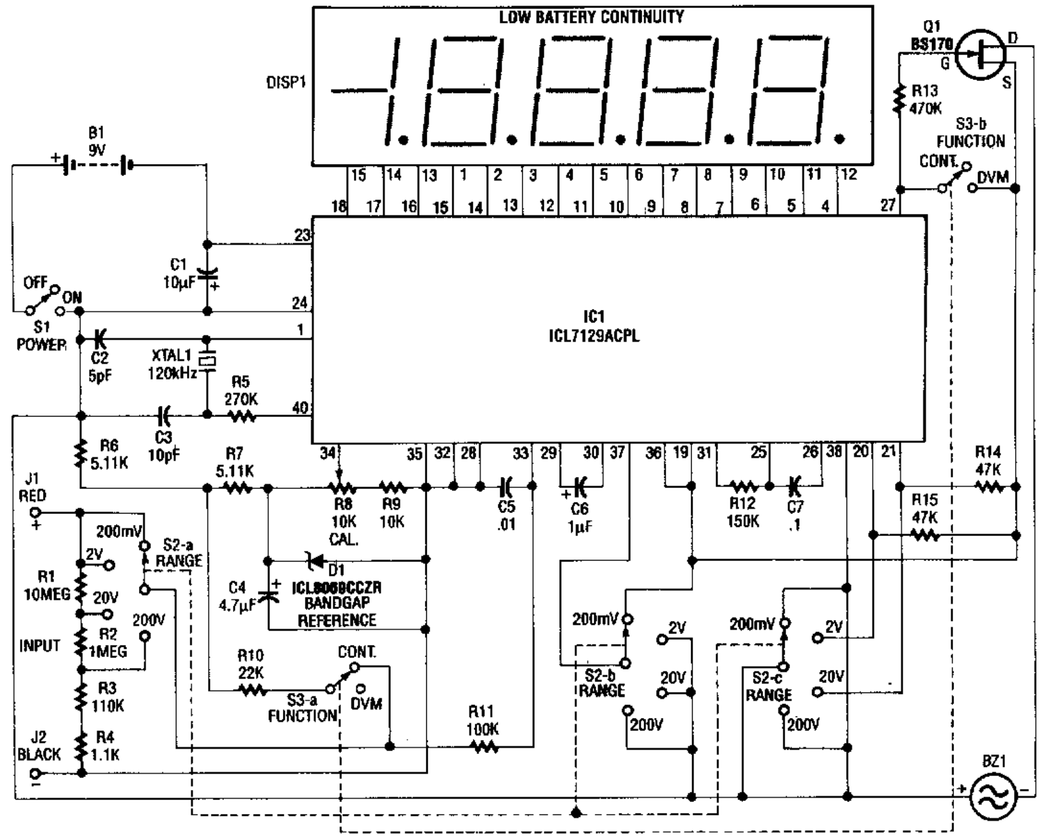
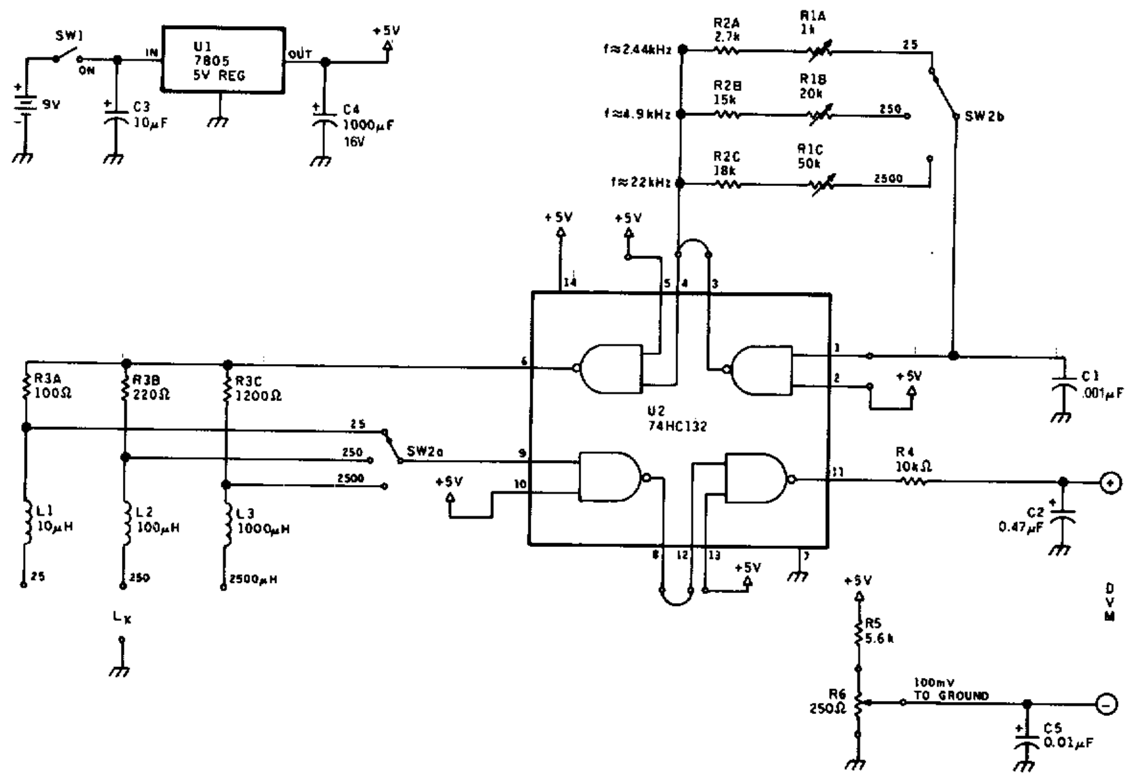


Fig. 50-2

This 4½-digit DVM circuit is built around a Maxim ICL7129ACPL A/D converter and LCD driver. An ICL8069 CCZR 1.2-V band-gap reference diode is used for a voltage reference. S2a-b-c select one of four ranges up to 200 V (maximum). The meter also has a piezoelectric buzzer for continuity testing. S3 selects either DVM or continuity. Crystal 1 can be changed to 100 kHz if maximum rejection of 50 Hz is desired. The crystal normally provides 120 kHz for best 60-Hz rejection. This is caused by the dual-slope conversion technique used in IC1.

INDUCTANCE WITH DVM MEASURING CIRCUIT

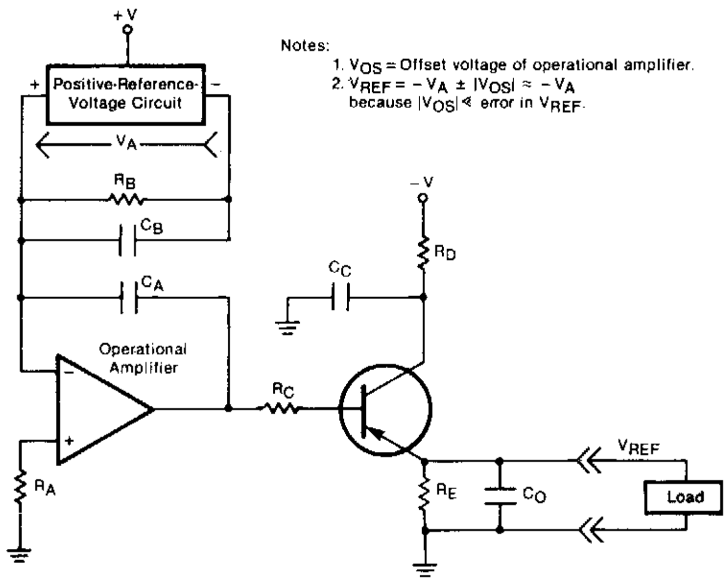


73 AMATEUR RADIO TODAY

Fig. 50-3

The inductance meter comprises an oscillator and pulse shaper. The square wave from the oscillator is differentiated and the differentiator output is shaped and read on a DVM. This will be proportional to inductance. R1A, B, C set calibration and R6 is an offset control.

NEGATIVE REFERENCE VOLTAGE CIRCUIT

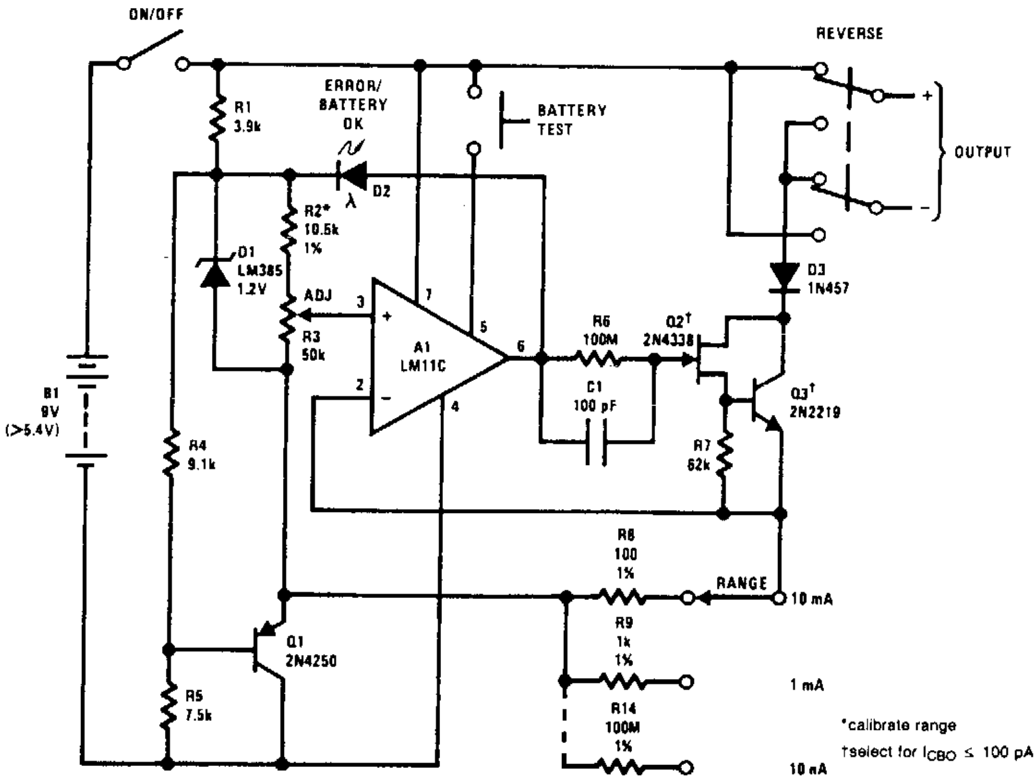


NASA TECH BRIEFS *Fig. 50-4*

This figure illustrates a circuit that supplies a precise negative reference voltage. To meet requirements of accuracy and stability, it incorporates a highly precise positive reference voltage by use of a high-gain, stable feedback booster circuit.

The booster circuit includes an operational amplifier and a transistor, which handles the load current. Typically, a positive-reference-voltage circuit can handle only relatively small load currents. This consideration does not apply in the present circuit because the positive-reference-voltage unit is placed in the voltage feedback loop of the booster circuit in parallel with resistor R_B. Thus, from the perspective of the positive-reference-voltage unit, R_B is a constant load. This feature enhances the stability of the circuit by removing the load regulation factor.

Provided that the offset voltage of the operational amplifier is low, the accuracy of the overall circuit depends only on the accuracy of the positive-reference-voltage unit. The overall circuit draws very little power for its own operation. It can handle unexpectedly heavy loads; the feedback configuration and the high gain provided by the combination of the operational amplifier and the transistor give the circuit a very low output impedance. The capacitors reduce the noise voltage and help stabilize the circuit. In the event that the load becomes a short circuit, R_D protects the transistor by limiting the load current.



NATIONAL SEMICONDUCTOR

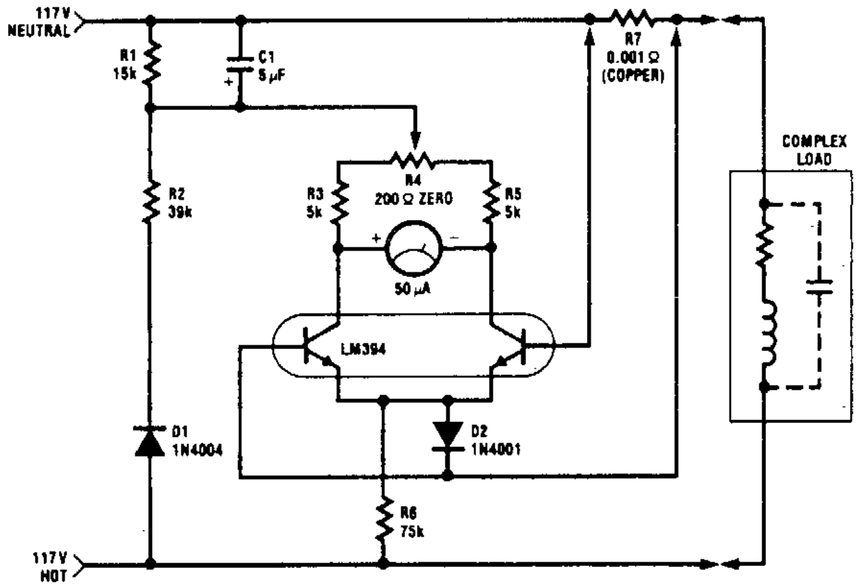
Fig. 50-5

A complete, battery-powered current source suitable for laboratory use is shown. The op amp regulates the voltage across the range resistors at a level determined by the voltage on the arm of the calibrated potentiometer, R3. The voltage on the range resistors is established by the current through Q2 and Q3, which is delivered to the output.

The reference diode, D1, determines basic accuracy. Q1 is included to ensure that the LM11 inputs are kept within the common-mode range with diminishing battery voltage. A light-emitting diode, D2, is used to indicate output saturation. However, this indication cannot be relied upon for output-current settings below about 20 nA, unless the value of R6 is increased. The reason is that very low currents can be supplied to the range resistors through R6 without developing enough voltage drop to turn on the diode.

If the LED illuminates with the output open, there is sufficient battery voltage to operate the circuit. But a battery test switch is also provided. It is connected to the base of the op-amp output stage and forces the output toward V+.

1-kW POWER METER

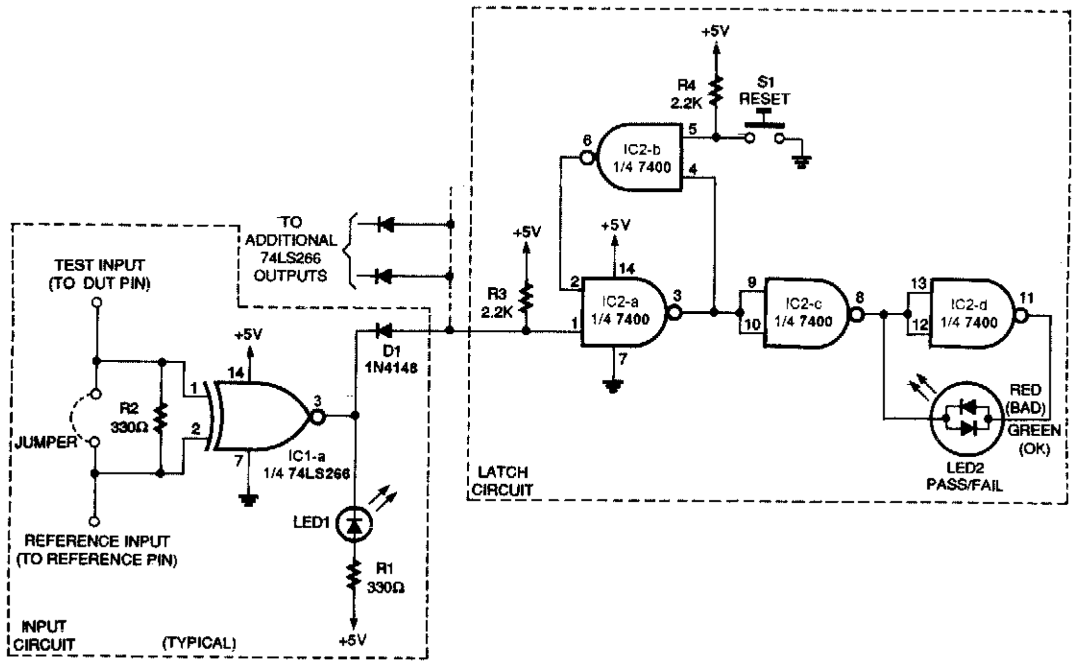


NATIONAL SEMICONDUCTOR

Fig. 50-6

The power meter shown uses only one transistor pair to provide the complete (*X*) (*Y*) function. The circuit is intended for 117 Vac ± 50 Vac operation, but can be easily modified for higher or lower voltages. It measures true (nonreactive) power being delivered to the load and requires no external power supply. Idling power drain is only 0.5 W. The load current-sensing voltage is only 10 mV, keeping load voltage loss to 0.01%. Rejection of reactive load currents is better than 100:1 for linear loads. Nonlinearity is about 1% full scale when using a 50-μA meter movement. The temperature correction for gain is accomplished by using a copper shunt (+0.32%/°C) for load-current sensing. This circuit measures power on negative cycles only, so it cannot be used on rectifying loads.

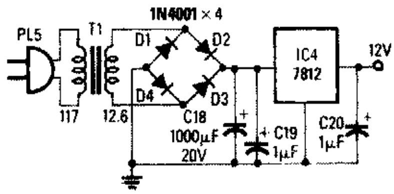
LOGIC CHIP TESTER



ELECTRONICS NOW *Fig. 50-7*

This device compares two logic signals and indicates whether the two logic signals are the same or different. To use the tester, first connect the test input to the suspect pin of the DUT. Next, connect the reference input to the same pin of an identical reference chip that is known to be good. Push the reset button to begin the test; the green section of the bicolor LED will be illuminated. Any signal on the test device that differs from the one on the reference device will then momentarily light the LED lamp that corresponds to that pin, and also latch on the red section of the bi-color LED. That indicates that the device under test is faulty. If the reference and DUT signals are the same, the DUT is OK, and the green LED will remain lit.

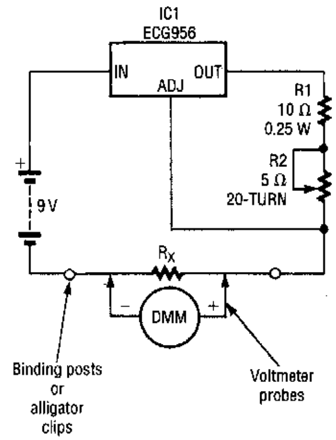
POWER SUPPLY FOR 10-MHz FREQUENCY STANDARD



RADIO-ELECTRONICS ***Fig. 50-8***

This simple power supply can be used in place of battery B1 of the 10-MHz frequency standard.

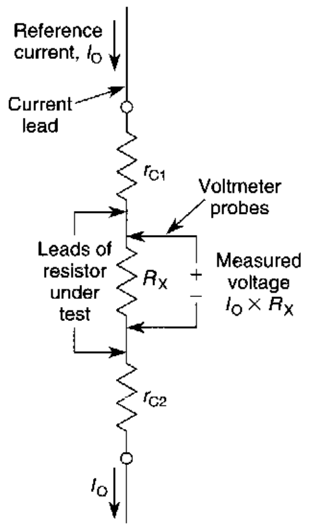
THREE-TERMINAL REGULATOR CURRENT SOURCE



ELECTRONICS NOW ***Fig. 50-9***

A three-terminal voltage regulator acts as a current source in this circuit. A resistor is being calibrated using a DMM and the current source.

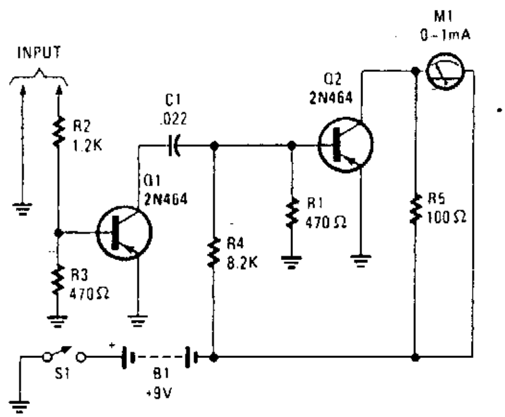
FOUR-WIRE RESISTANCE MEASUREMENT HOOKUP



ELECTRONICS NOW ***Fig. 50-10***

A true four-wire resistance measurement hookup.

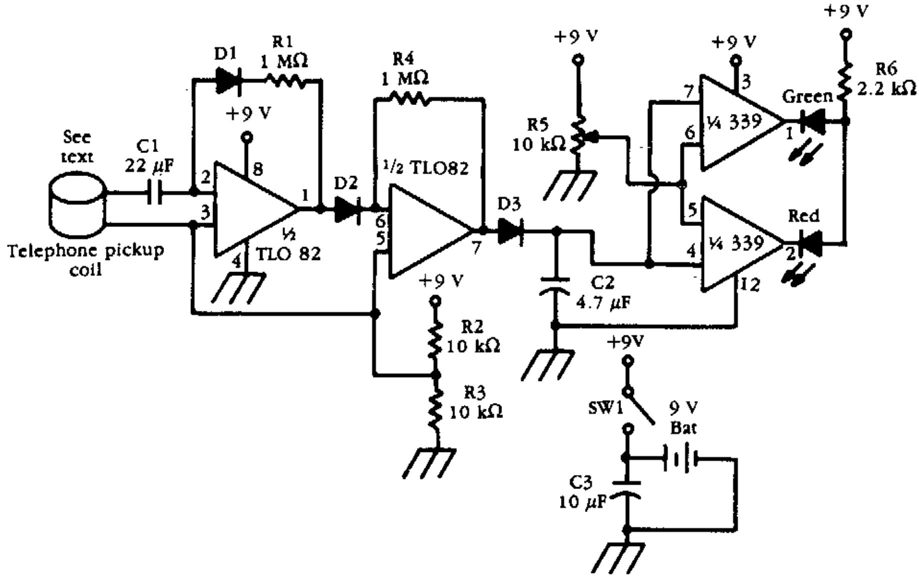
AUDIO FREQUENCY METER



POPULAR ELECTRONICS ***Fig. 50-11***

A pulse-shaper is used in a tachometer circuit to drive a meter.

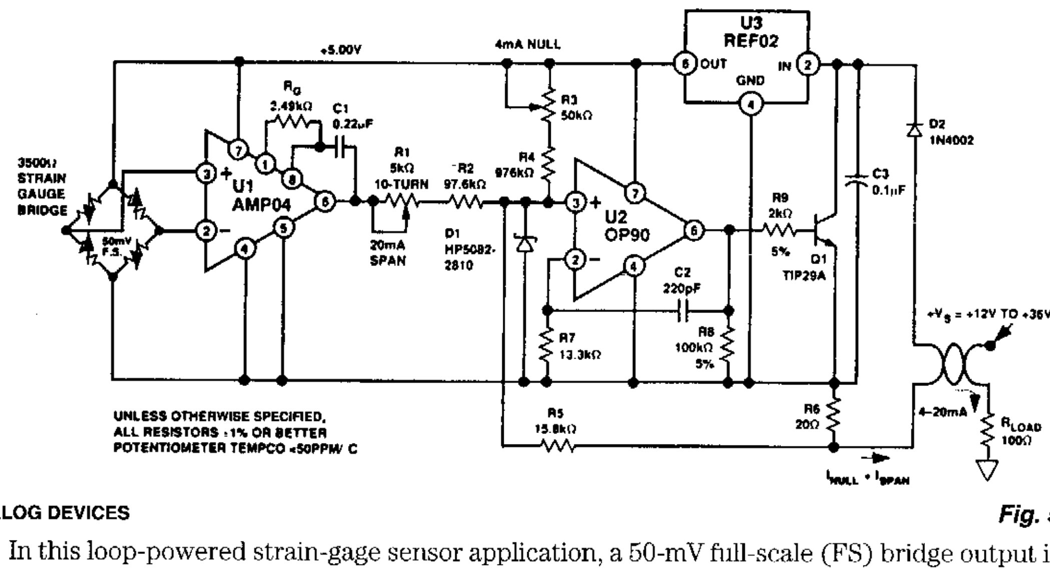
ELF MONITOR



McGRAW-HILL *Fig. 50-12*

A telephone pick-up coil is used as a sensor for low-frequency magnetic fields. The signal is amplified and detected, then used to drive a comparator.

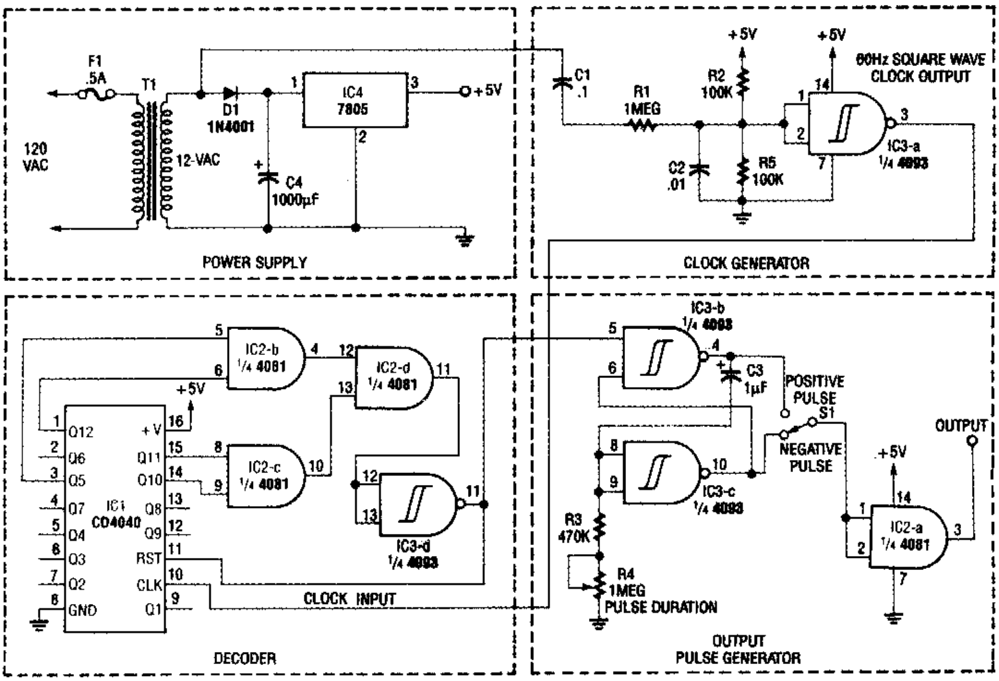
STRAIN-GAGE SENSOR



ANALOG DEVICES *Fig. 50-13*

In this loop-powered strain-gage sensor application, a 50-mV full-scale (FS) bridge output is amplified and calibrated for a 4–20-mA transmitter output. Power is furnished by the remote loop supply of 12 to 36 V.

MINUTE MARKER



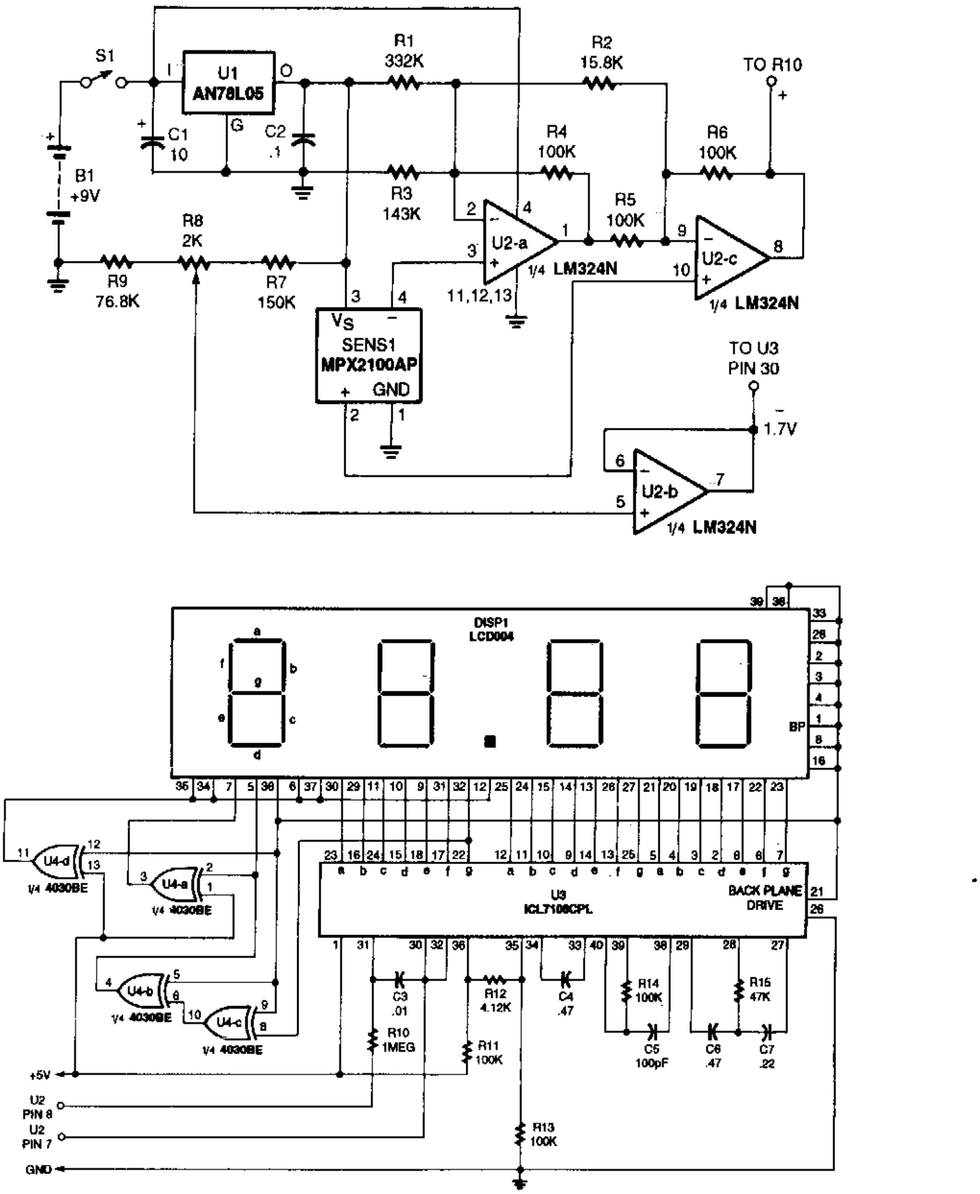
ELECTRONICS NOW *Fig. 50-14*

The figure shows the schematic of a minute marker. The output of transformer T1 is 12.6 Vac at 60 Hz, which is rectified by D1 and regulated by IC4, and LM7805 regulator, to provide 5 Vdc for the circuit. The unrectified ac is bandpass-filtered by R1, R2, R5, C1, and C2. Resistors R2 and R5 also form a dc-voltage divider, which biases the input of Schmitt trigger IC3-a to 2.5 V. The Schmitt trigger generates a 60-Hz square wave, which is fed to the input of IC1, a CE4040 12-stage binary counter.

The outputs of the counter are a 4081 quad AND gate (IC2), and the decoded output is fed back to the reset input of the counter, which resets the counter when the desired count is reached.

The pulse from IC2-d is inverted by Schmitt trigger IC3-d, and passed along to the output pulse generator. The output pulse is generated by two Schmitt triggers cross-connected as an RS flip-flop (IC3-b and IC3-c). The output of the flip-flop is fed to 3, R4, and C3, whose values set the output pulse duration. The output pulse duration (T) can be approximated by the formula $T = 1.2 \times C_3 \times (R_3 + R_4)$. A positive or negative-going pulse is selected by S1, and buffered by the remaining AND gate (IC2-a).

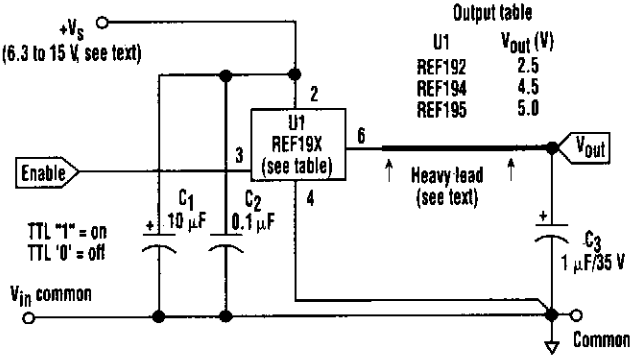
DIGITAL BAROMETER



POPULAR ELECTRONICS ***Fig. 50-15***

A pressure sensor is used in this application. This outputs a voltage to amplifier U2, and a 3½ digit A/D converter module. It is calibrated to read barometric pressure in inches of mercury.

REFERENCE CIRCUIT

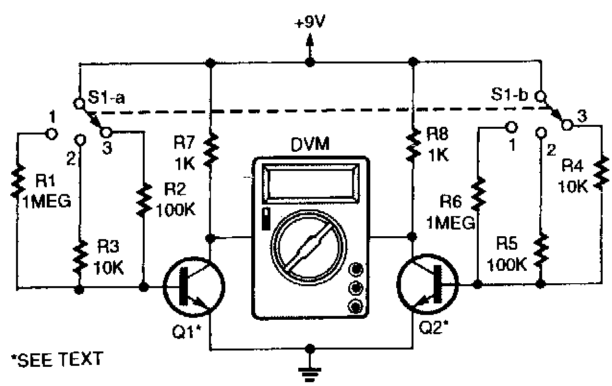


ELECTRONICS DESIGN

Fig. 50-16

In this high-performance reference circuit, U1 is a device from the REF190 series producing device-selectable outputs of 2.5, 4.5, and 5 V with simple, noncritical external circuitry. An Analog Devices REF 19 X (see the table in the figure) is used to derive a reference voltage.

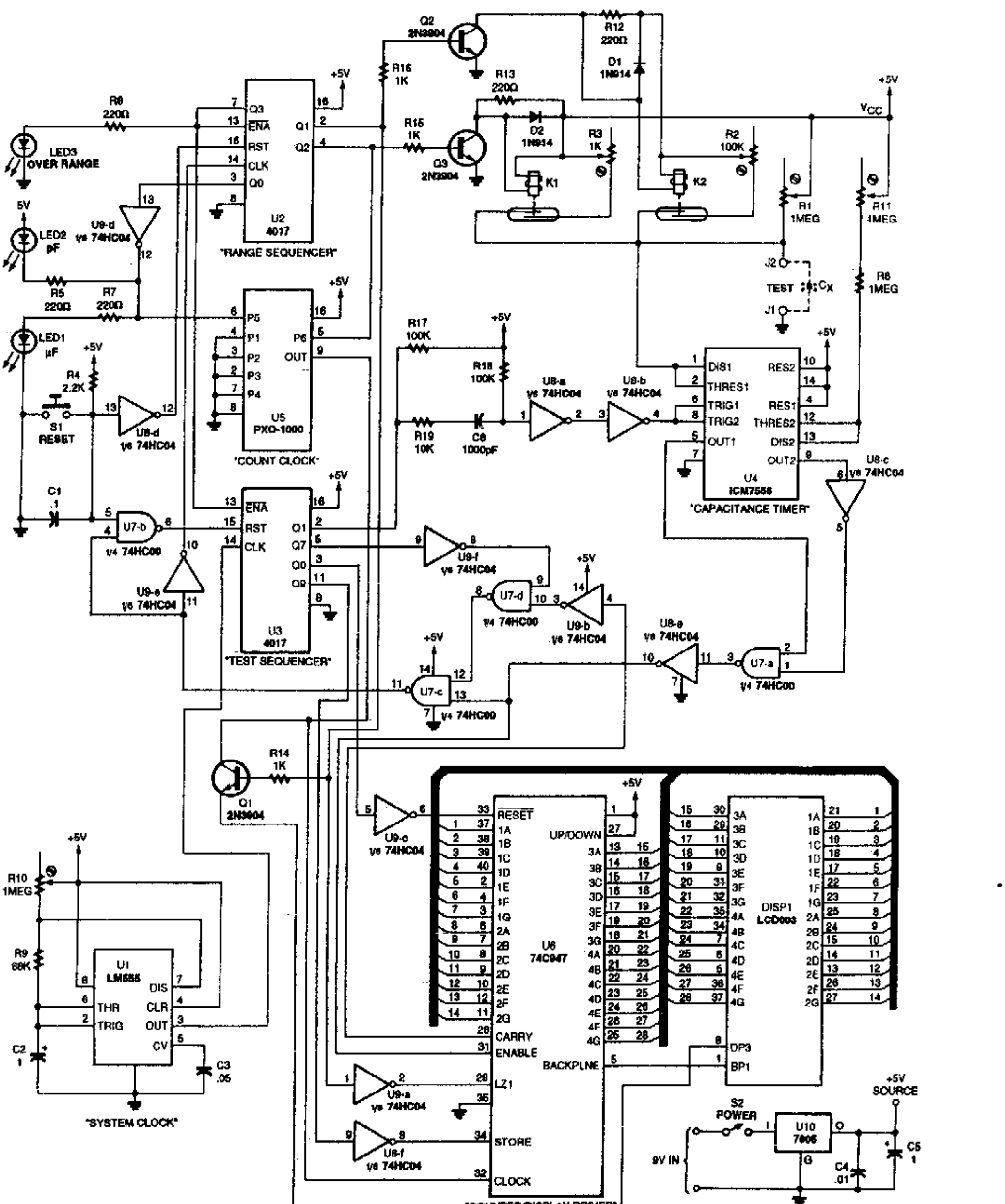
TRANSISTOR MATCHING CIRCUIT



POPULAR ELECTRONICS

Fig. 50-17

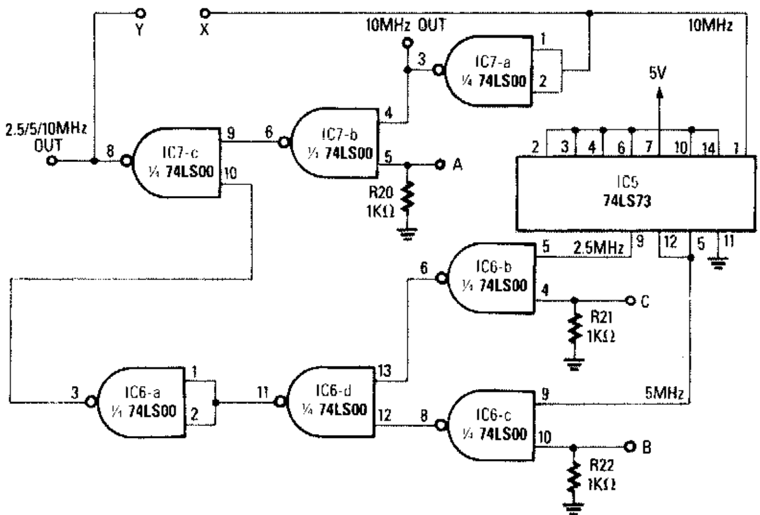
In order to match two transistors, select Q1 and Q2 so that they give zero readings (or as close as possible) on a DVM. The DVM acts as a null detector. An analog meter can be substituted. S1 should be set for an appropriate level of base current (approximately 8, 80, or 800 µA).



POPULAR ELECTRONICS *Fig. 50-18*

This digital capacitance meter reads from 1 pF to 1000 µF. Basically, a timer (U4) uses the unknown capacitance to generate a pulse of duration, depending on the value of unknown capacitance, and the pulse duration is measured. The display is an LCD 0003 driven by a 74C947 counter/display driver.

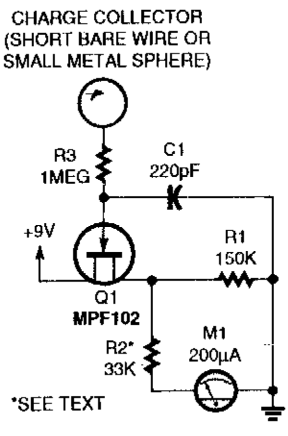
FREQUENCY DIVIDER FOR 10-MHz FREQUENCY STANDARD



RADIO-ELECTRONICS

Fig. 50-19

ELECTROSCOPE



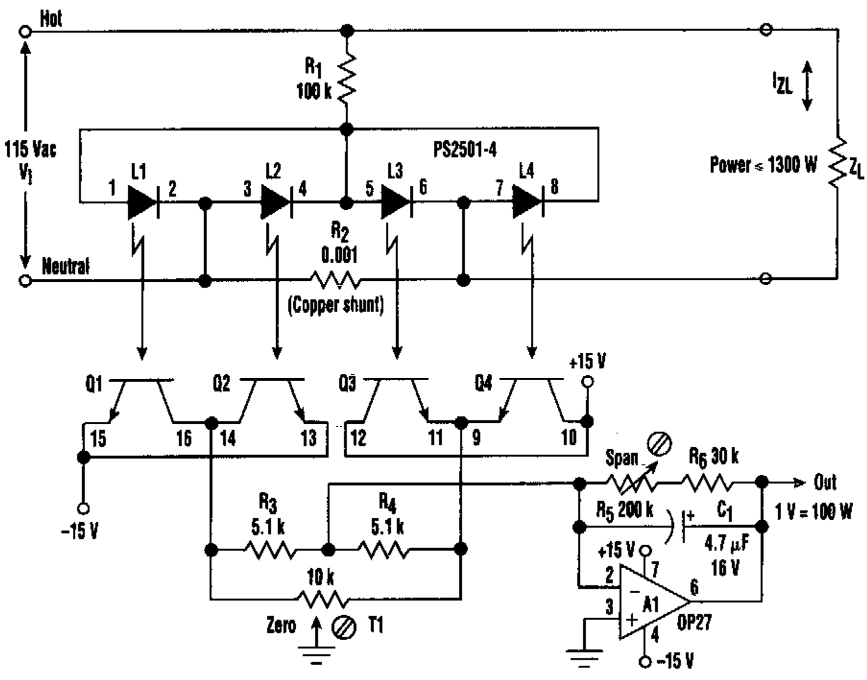
This circuit is useful for detecting electrostatic charges. In operation, C1 reduces ac noise, but lowers the sensitivity a bit. The MPF102 and R1 form a voltage divider. When the FET's gate is earth-grounded, the divider's output will be about 4.5 V giving a half-scale reading on M1, a 200-μA meter. A positively charged object (like cotton-rubbed glass) will give a positive deflection from half-scale, and a negatively charged object (a plastic comb, for example) will give a negative meter deflection.

The whole circuit (including the 9-V battery supply) should be in a metal enclosure, and a short piece of bare wire makes a fine charge collector.

POPULAR ELECTRONICS

Fig. 50-20

OPTICAL ISOLATOR WATTMETER

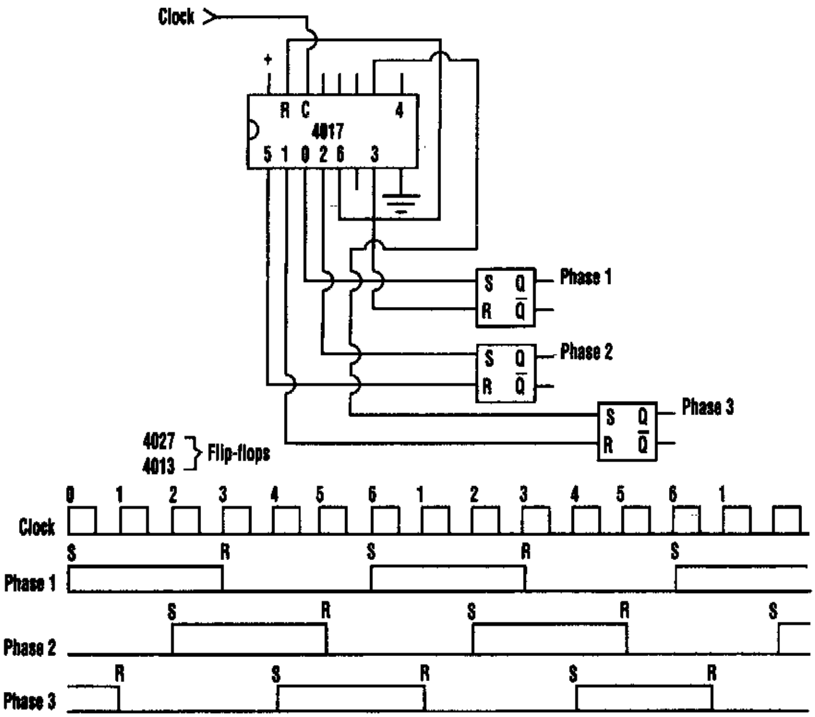


ELECTRONIC DESIGN

Fig. 50-21

The quad-channel optical isolator, consisting of LED L1 through L4 and phototransistors Q1 through Q4, is connected in a double bridge configuration. The arrangement serves to compute the four-quadrant product of ac line voltage and Z1 load current. The result is an accurate representation of the true instantaneous power delivered to the load—even if the line voltage wanders and the load is reactive and nonlinear. This wattmeter function is, of course, optically isolated from the ac line, has full-scale limit of 1300 W, and is output with scale factor of 1 V/100 W.

DIGITAL THREE-PHASE WAVE GENERATOR

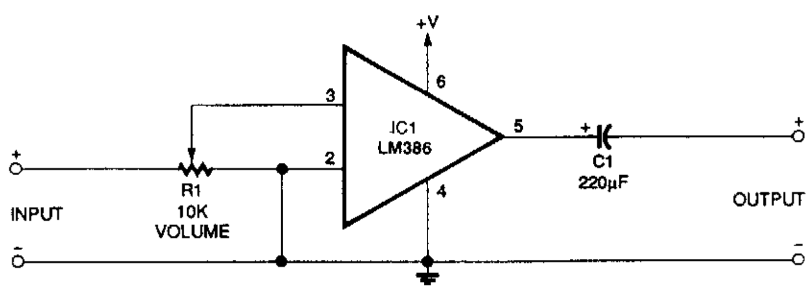


ELECTRONIC DESIGN **_Fig. 50-22_**

With a simple digital circuit, three-phase square waves can be produced from a single-phase square-wave signal source. The timing diagram shows that the second and third phases are 120° and 240° behind the first phase, respectively.

The frequency range over which the three-phase outputs will occur is limited only by the capability of the logic used. The output frequency is ⅙ of the input frequency.

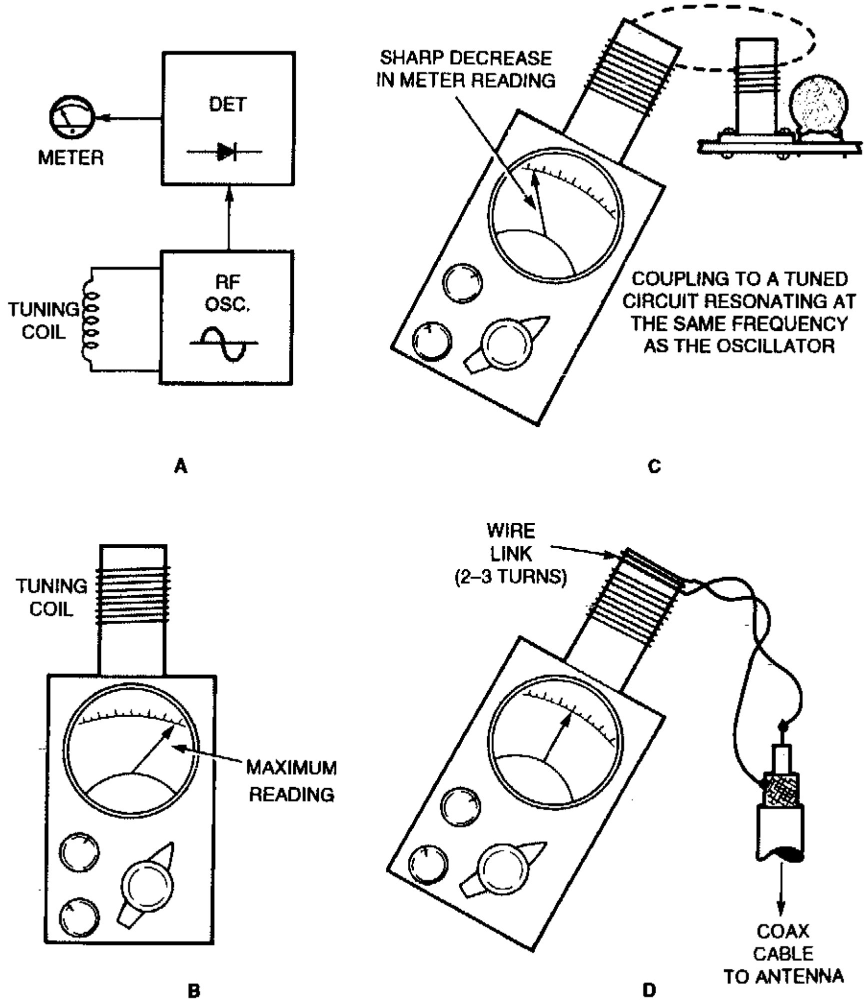
SIMPLE TEST AUDIO AMPLIFIER



ELECTRONICS NOW **_Fig. 50-23_**

This circuit has a gain of about 20. A suitable power supply voltage is 5 to 12 V, depending on the desired audio output power level.

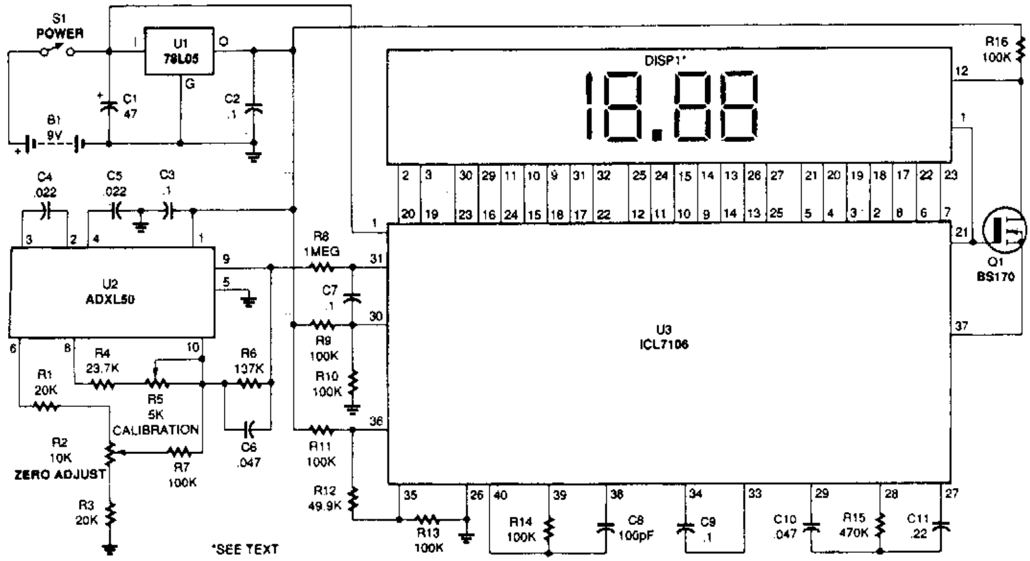
GATE DIP OSCILLATOR I



POPULAR ELECTRONICS *Fig. 50-24*

The typical dip meter is comprised of a tuning coil, RF oscillator, a detector, and a meter as shown in A. When the meter's tuning coil is coupled to a tuned circuit resonating at the same frequency as the GDM, the reading dips (C). The GDM's tuning coil can be coupled to the coaxial feed line of an antenna through a few (perhaps 2 to 3) turns of wire, and used to determine the antenna's resonant frequency (D).

ACCELEROMETER (G METER) CIRCUIT



POPULAR ELECTRONICS ***Fig. 50-25***

As this schematic shows, the ADXL50 accelerometer, U2, interfaces with an A/D converter, U3, to drive a 3½-digit LCD module, DISP1. Because that module displays any number from –19.99 to +19.99, the circuit is designed to measure g's within that range. The heart of the circuit is U2, the ADXL50 accelerometer. The sensitivity of that chip is set to ±20 g's in order to accommodate the full scale capability of LCD module DISP1 (19.99). Circuit gain is determined by the values of R4, R5, and R6, potentiometer R2, and R3 provides a way to manually set the zero-g voltage-output level at pin 9 of U2 to half the supply voltage—2.5 V. That output voltage will vary linearly by 0.1-V/g of acceleration.

In order to achieve good circuit performance at low g levels, the bandwidth of the amplifier is limited to about 30 Hz by C6. The digital-display section of the circuit is composed of DISP1 and U3. Included in U3 are the A/D converter, clock oscillator, storage resistors and latches, 3½-digit seven-segment decoders, and backplane generator.

The differential analog input of U3 is applied between pins 30 and 31. The positive input, pin 31, is driven by output-pin 9 of U2 through R8, a buffer resistor, and the negative input, pin 30, is biased at a fixed voltage of 2.5 V by a voltage-divider string composed of R9 and R10.

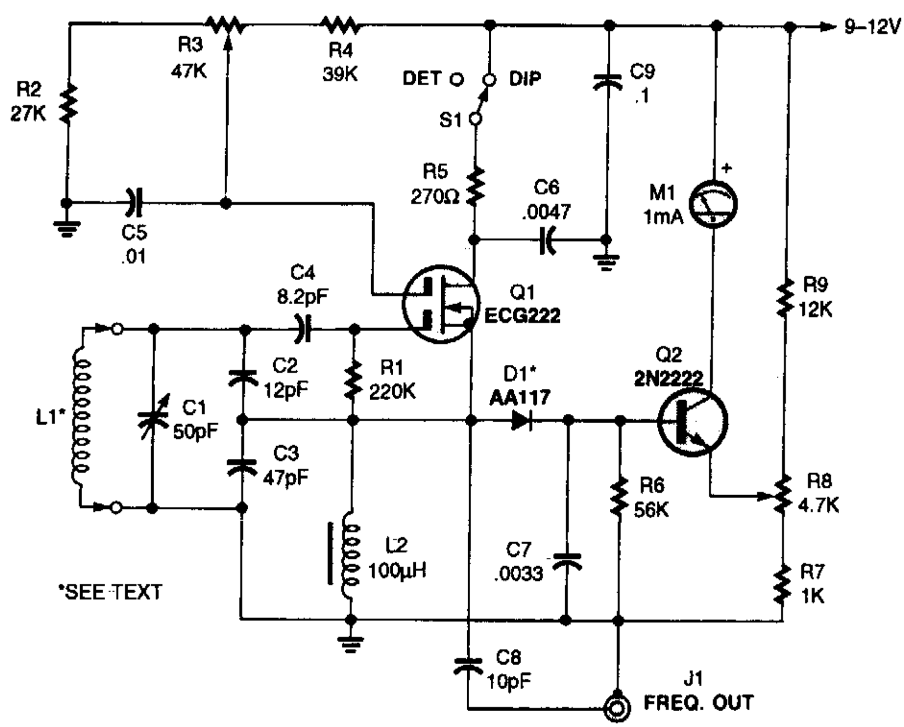
A reference voltage is required by U3. Full-scale display, 19.99, occurs when the differential, analog input voltage applied between pins 31 and 30 is equal to twice the reference voltage. The decimal point of the LCD has to be illuminated to display readings from 0.00 to 19.99. That is done by inverting the backplane square-wave drive signal appearing at pin 21 of U3, through MOSFET Q1, and applying the 180-degree out-of-phase signal to pin 12 of DISP1.

GATE DIP OSCILLATOR II

TABLE 1—COIL WINDING DATA

Band (MHz)	Turns	Wire Size/Type
3.5 6.5	45	32-enameled
6.5 11	32	26-enameled
11 19	14	20-enameled
15 24	10	20-enameled
21 36	7	insulated connection wire
32 56	4	insulated connection wire
60 .110	U-shaped*	16 enameled

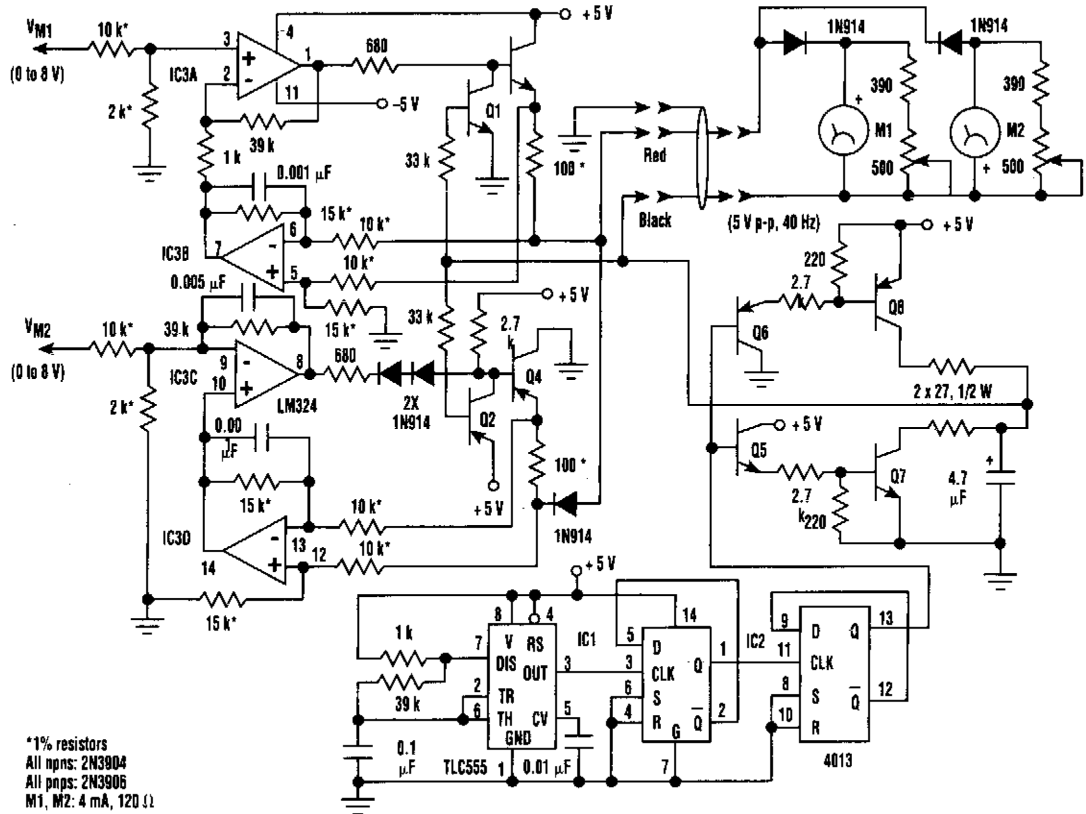
*1.8-inches long



POPULAR ELECTRONICS *Fig. 50-26*

Useful for measuring the resonant frequencies of antennas, tuned circuits, and also as a tuned detector, this circuit is a modern variation of the classic vacuum tube grid dip oscillator. It coupled the G.D.O. to a tuned circuit and caused RF energy to be absorbed by the unknown tuned circuit when the G.D.O. frequency was the same as that of the tuned circuit in question. This showed as a "dip" in the meter reading.

TWO REMOTE METERS



ELECTRONIC DESIGN

Fig. 50-27

Two remote meters can be driven independently using just one wire pair. This "constant current" design eliminates the effects of wire-pair resistance up to 200 Ω. Driving two remote meters independently usually requires two wire pairs (one pair for each meter).

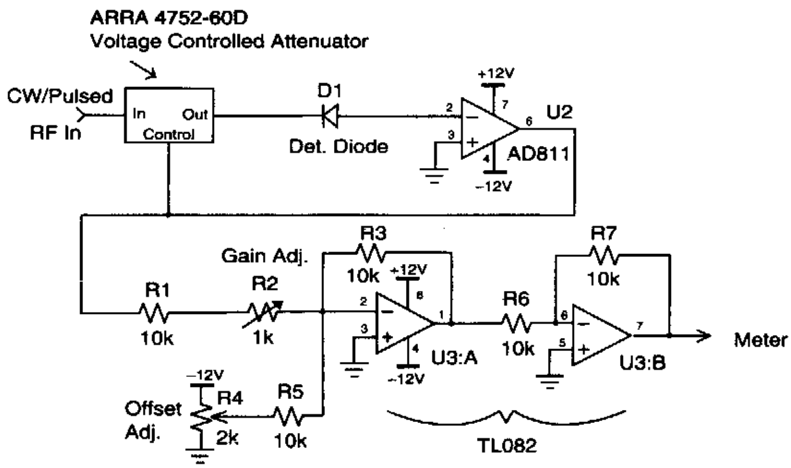
In the circuit, IC1 and IC2 generate a 40-Hz symmetrical square wave (the frequency isn't critical). Q5 through Q8 amplify the square wave to 5 V p-p, which is applied to the "return" (black wire) for the remote meters.

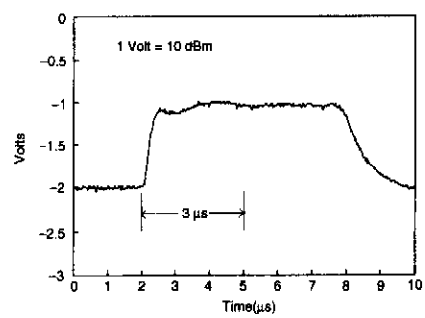
Amplifier IC3A buffers the input signal voltage V_{mi}, intended for meter M1 (0 to 8 V), and sends it through emitter-follower Q3 to a 100-Ω current-sense resistor. The other end of this resistor is tied to the "supply" (red wire) of the remote meters. IC3B amplifies the voltage across the sense resistor, which corresponds to the current sent to remote meter M1, and closes the feedback loop to IC3A.

This results in a voltage of 0 to 8 V at the M1 input, generating a current of 0 to 10 mA to M1. Transistor Q1 gates this current on and off synchronous to the 40-Hz square wave so that meter M1 actually sees a 50% 0-to-+10-mA peak (0 to 5 mA average) current.

Similarly, IC3C, IC3D, Q2, and Q4 provide a 0- to −10-mA peak current for M2. M1 and M2 are isolated by the two-reverse-connected 1N914 diodes in the remote-meter box. Variable resistors across M1 and M2 permit calibration. The extra 1N914 diode in the M2 drive circuit prevents interference between M1 and M2.

NOVEL RF POWER METER





RF DESIGN

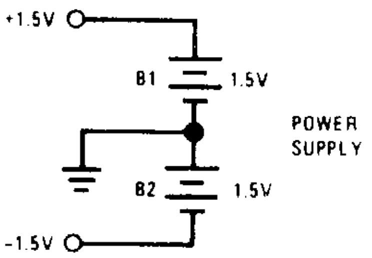
Fig. 50-28

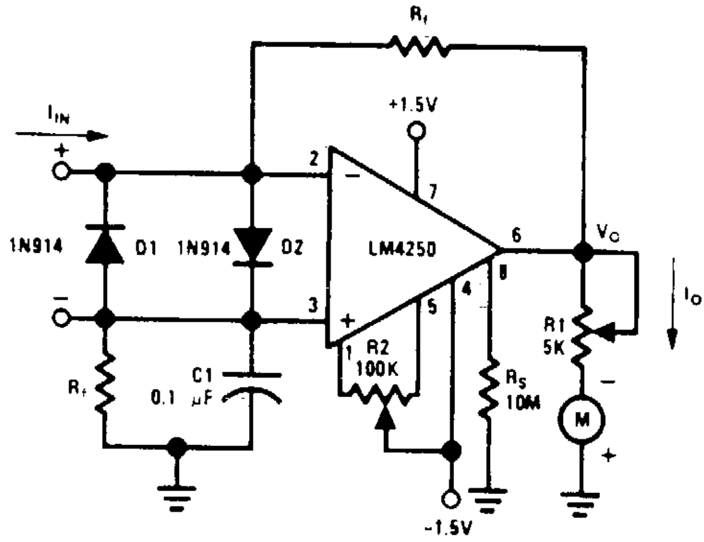
The circuit matches the diode with a voltage-variable attenuator that has a logarithmic response. By varying the attenuation until the diode output is zero, the resulting attenuation value then corresponds to the input power level. Because the voltage-variable attenuator's output is logarithmic, diode nonlinearities become negligible.

NANOAMMETER

Resistance Values for DC Nano and Micro Ammeter

I FULL SCALE	R_f [Ω]	R'_f [Ω]
100 nA	1.5M	1.5M
500 nA	300k	300k
1 µA	300k	0
5 µA	60k	0
10 µA	30k	0
50 µA	6k	0
100 µA	3k	0





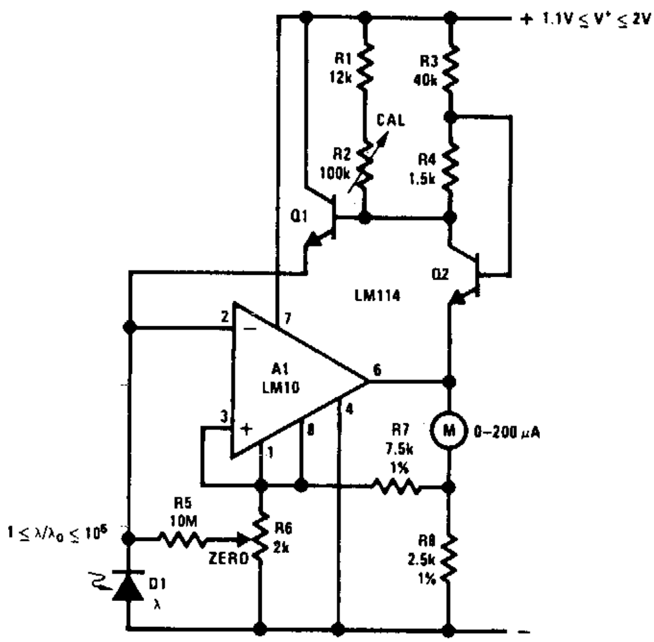
NATIONAL SEMICONDUCTOR

Fig. 50-29

Potentiometer R2 provides an electrical meter zero by forcing input offset voltage V_{os} to zero. Full-scale meter deflection is set by R1. Both R1 and R2 only need to be set once for each op amp and meter combination. For a 50-µA 2-kΩ meter movement, R1 should be about 4 kΩ to give full-scale meter deflection in response to a 300-mV output voltage. Diodes D1 and D2 provide full input protection for overcurrents up to 75 mA.

With an R_f resistor value of 1.5 MΩ, the circuit becomes a nanoammeter with a full-scale reading capability of 100 nA. Reducing R_f to 3 kΩ in steps, as shown in the figure increases the full-scale deflection to 100 µA, the maximum for this circuit configuration. The voltage drop across the two input terminals is equal to the output voltage (V_o) divided by the open loop gain. Assume that an open loop gain of 10,000 gives an input voltage drop of 30 µV or less.

1.5-V LOGARITHMIC LIGHT LEVEL METER



NATIONAL SEMICONDUCTOR

Fig. 50-30

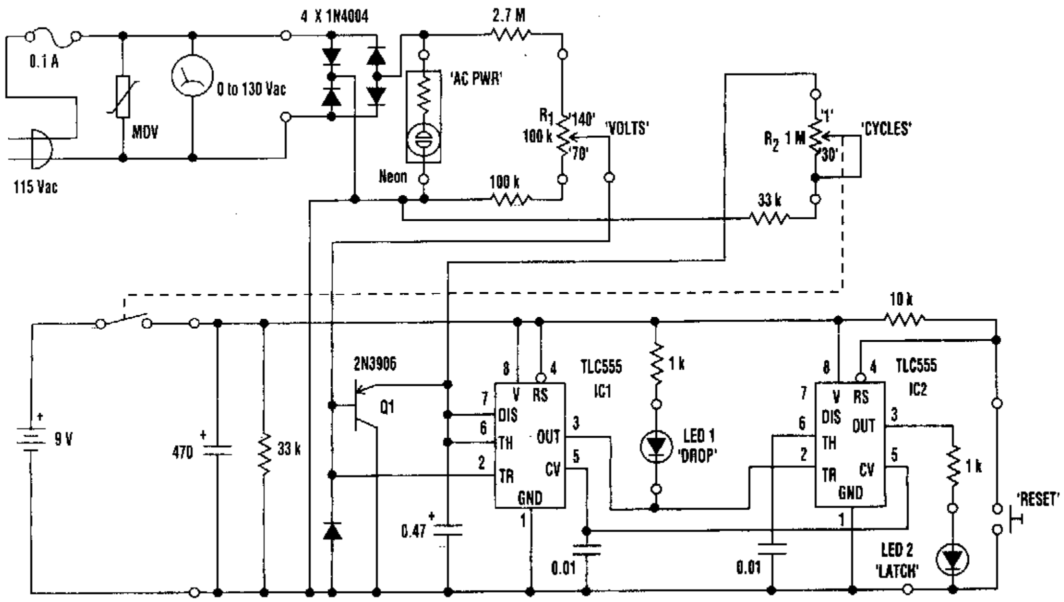
A portable light-level meter with a five-decade dynamic range is shown. The circuit is calibrated at mid-range with the appropriate illumination by adjusting R2 such that the amplifier output equals the reference and the meter is at center scale. The emitter-base voltage of Q22 will vary with supply voltage; so R4 is included to minimize the effect on circuit balance. If photocurrents less than 50 nA are to be measured, it is necessary to compensate the bias current of the op amp.

The logging slope is not temperature compensated. With a five-decade response, the error at the scale extremes will be about 40% (a half stop in photography) for a ±18°C temperature change.

If temperature compensation is desired, it is best to use a center-zero meter to introduce the offset, rather than the reference compensation. It can be obtained by making the resistor in series with the meter a copper wire-wound unit.

If this design is to be used for photography, it is important to remember that silicon photodiodes are sensitive to near-infrared light, whereas ordinary film is not. Therefore, an infrared-stop filter is called for. A blue-enhanced photodiode or an appropriate correction filter would also produce best results.

ac POWER MONITOR



All capacitors in microfarads.
Diodes are 1N914 except where otherwise noted.

ELECTRONIC DESIGN

Fig. 50-31

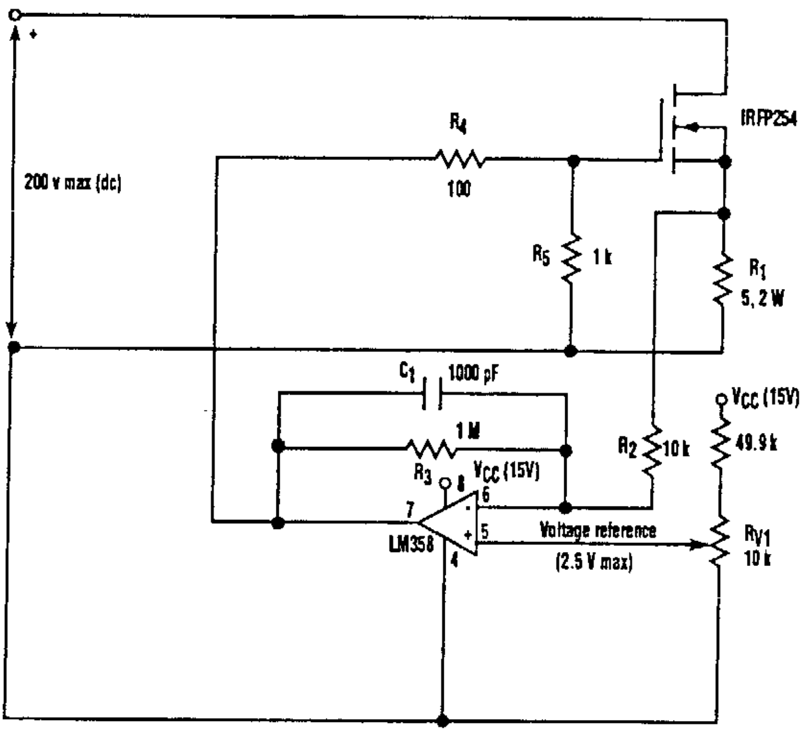
The 0- to 130-V voltmeter and neon "AC PWR" lamp provide an average indication of the ac power. The fuse and metal oxide varistor (MOV) protect the monitor against overvoltage spikes.

Four 1N4004 diodes rectify the ac voltage, generating negative-going pulses twice per cycle (every 8.33 ms for 60-Hz power). Variable-resistor F1 supplies a reduced amplitude sample of these pulses to a missing pulse detector consisting of Q1, IC1, and associated circuitry. As long as the pulse amplitude exceeds the threshold value set by R1, IC1 continually triggers, keeping its output high. When the pulse amplitude drops below the threshold value, IC1 times out with a time constant set by variable resistor R2 and the 0.47-μF capacitor. R2 is calibrated to read the number of cycles required for a dropout indication. It can be set between 1 cycle (about 17 ms) and 30 cycles (0.5 second).

When IC1 times out, its output goes low. This turns on LED1. The low output also triggers IC2, which is configured as a set-reset flip-flop. This turns on LED2. When the voltage returns to normal, IC1 again starts triggering and its output returns high, turning off LED1. LED2, however, remains on until the manual reset button is pressed.

The circuit is powered by a 9-V battery and is assembled in a plastic or grounded metal case. Notice that there's no isolation between the ac power line and the monitor circuitry. Be careful to avoid electrical shock when testing the circuitry.

100-W VARIABLE RESISTOR SIMULATOR



ELECTRONIC DESIGN

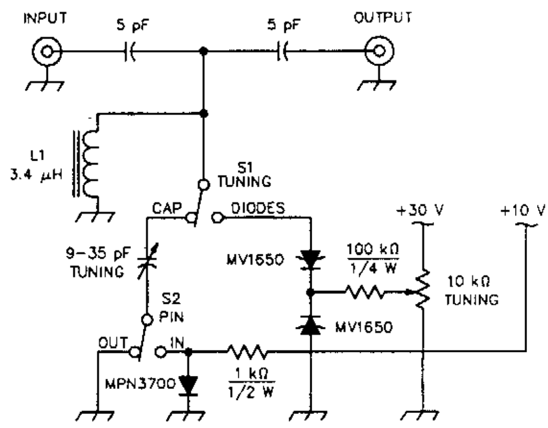
Fig. 50-32

Variable resistive loads with precise load steps often are required in automatic testers used to test and calibrate power supplies. The action of a high-power (100 W) variable resistor can be simulated with the circuit shown.

The voltage drop across R1, which is proportional to the FET current, is compared against a variable input voltage reference using a high-gain op amp. Error voltage developed by the amplifier drives the gate, controlling the transconductance of the FET.

Power dissipation is limited by the safe-operating-area curve of the selected FET. The FET should be mounted onto a properly sized heatsink or a heatsink-fan combination to maintain its case temperature within safe limits. The circuit is designed to dissipate a maximum power of 100 W if the FET-case temperature is maintained below 50°C. The potentiometer (RV1) can be replaced by a digital-to-analog converter so that it can adapt to the computer control for use in automatic testers.

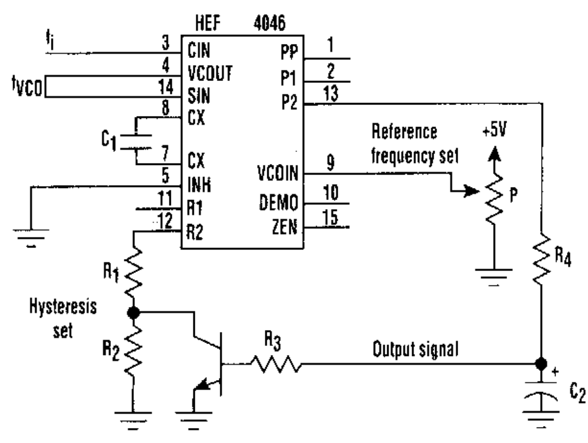
IMD TEST CIRCUIT FOR PIN DIODES



QST *Fig. 50-33*

A loosely coupled tuned circuit for testing IMD production by PIN and tuning diodes in a narrow-band preselector, S1, TUNING, selects whether C1 or a pair of back-to-back MV1650 tuning diodes resonate L1. S2, PIN, adds or removes an MPN3700 PIN diode in series with C1. L1 consists of 33 turns of #28 enameled wire on a t-37-6 toroidal powdered-iron core. The MV1650, a "20-V" tuning diode, exhibits a nominal capacitance of 100 pF at a tuning voltage of 4 V.

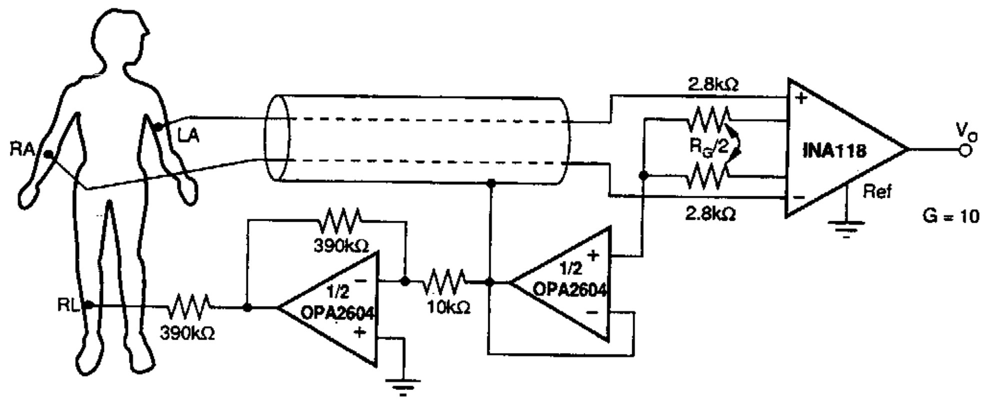
VCO AND INPUT FREQUENCY COMPARER



ELECTRONIC DESIGN *Fig. 50-34*

Comparison of an input signal's frequency (f_i) with that of voltage-controlled oscillator (f_{VCO}) can be accomplished with just one CMOS phase-locked loop IC and a transistor (see figure). The phase and the frequency can be compared with a phase comparator, which, along with the VCO, is part of the HEF4046 PLL IC. The transistor helps introduce hysteresis, enabling the circuit to be used as a switch driver.

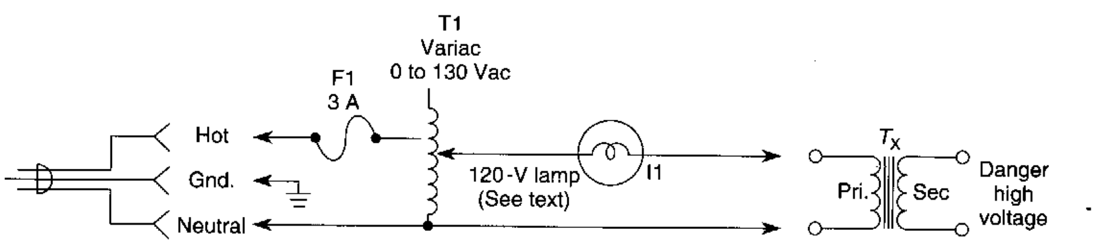
ECG AMPLIFIER WITH RIGHT LEG DRIVE



BURR-BROWN

Fig. 50-35

POWER TRANSFORMER TESTER



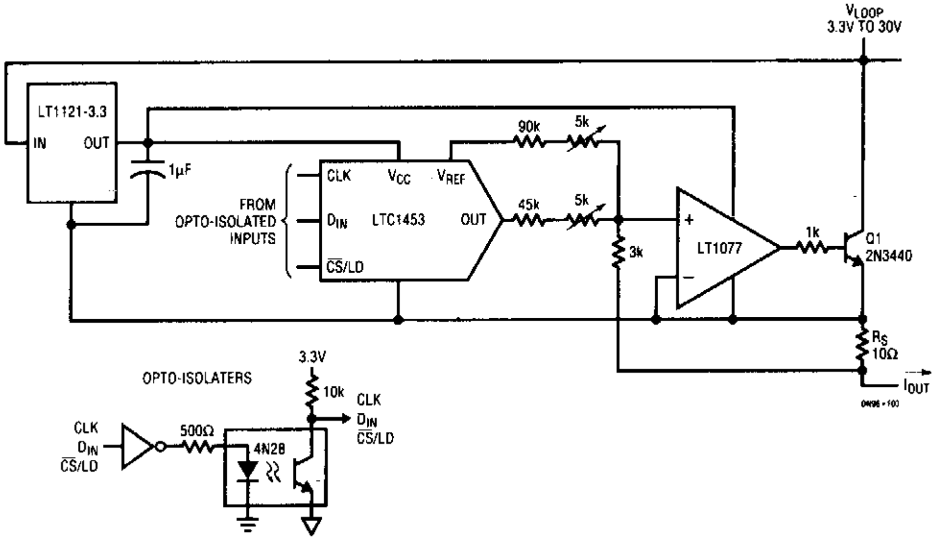
Warning: observe existing shock hazard

WILLIAM SHEETS

Fig. 50-36

Very often a power transformer is suspect and connecting a shorted transformer to an ac source can be hazardous. This test method will detect a defective or shorted transformer. The primary of the power transformer is energized through a Variac (0 to 130 Vac) and a lamp equal in wattage to about half that of the transformer under test. Connect the transformer, set Variac at zero, then energize circuit. Apply voltage to suspected transformer (Tx) as shown. The lamp should not light. If it does, Tx is shorted. Next, short the secondary of suspected Tx. This time, the lamp should light. For multiple winding transformers, repeat for each secondary winding. Beware of the shock hazard as the open windings of Tx can develop full-rated voltage.

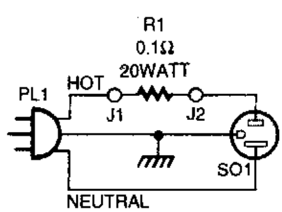
4- TO 20-mA PROCESS CONTROLLER



LINEAR TECHNOLOGY ***Fig. 50-37***

The figure shows how to use an LTC1453 to make an optoisolated digitally controlled 4- to 20-mA process controller. The controller circuitry, including the optoisolator, is powered by the loop voltage that can have a wide range of 3.3 V to 30 V. The 1.22-V reference output of the LTC1453 is used for the 4-mA offset current and V_{OUT} is used for the digitally controlled 0- to 16-mA current. R_S is a sense resistor and the LT1077 op amp modulates the transistor Q1 to provide the 4- to 20-mA current through this resistor. The control circuitry consumes well under the 4-mA budget at zero scale.

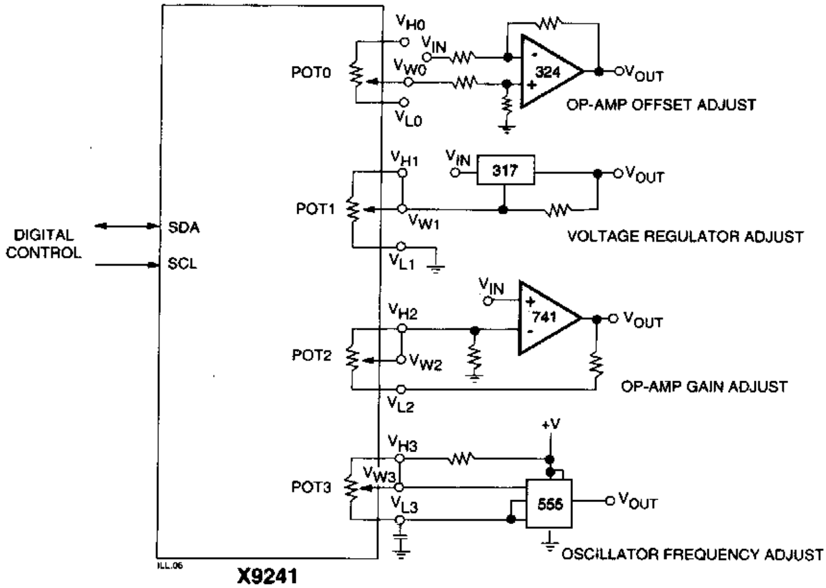
SIMPLE HIGH-CURRENT MEASURER



POPULAR ELECTRONICS ***Fig. 50-38***

Testing heavy-load devices with a ten-amp maximum meter can be accomplished with this straightforward meter add-on. If done right, it could be made from a high-current extension cord. J1 and J2 are well-insulated jacks to accept meter probe tips.

ANALOG CIRCUITRY CALIBRATOR

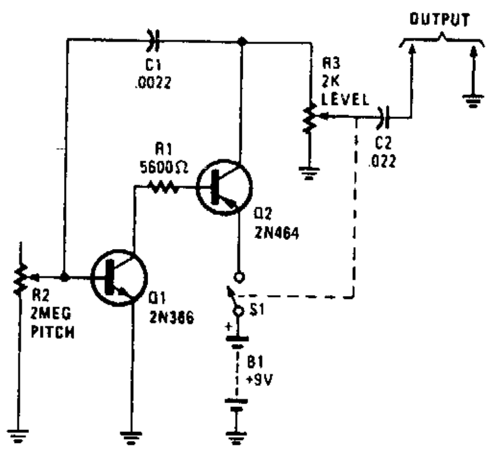


XICOR

Fig. 50-39

An XICOR X9241 Quad POT IC can be used to digitally adjust four analog circuits, as shown in the example schematic.

SIMPLE SIGNAL GENERATOR FOR SIGNAL TRACING

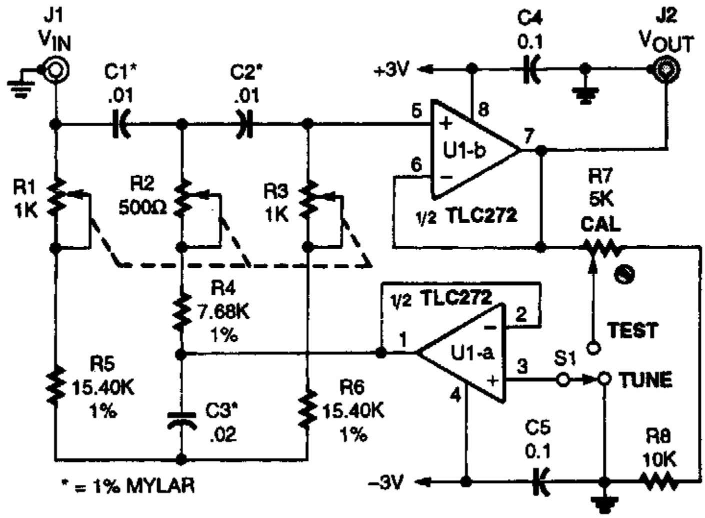


POPULAR ELECTRONICS

Fig. 50-40

A simple R-C oscillator generates a harmonic-rich waveform for signal injection.

SIMPLE HARMONIC DISTORTION ANALYZER



POPULAR ELECTRONICS

Fig. 50-41

This simple circuit lets you accurately measure the total harmonic distortion (THD) using your true-RMS voltmeter.

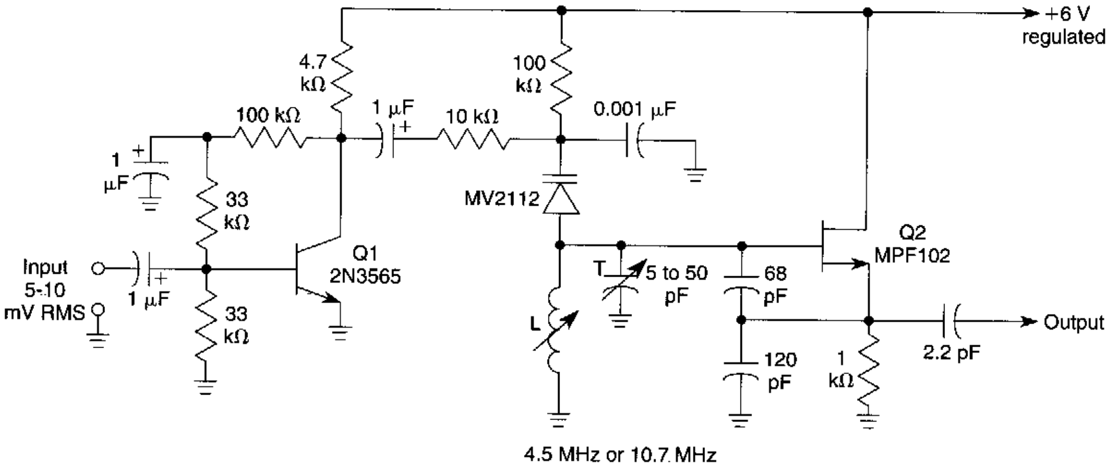
This THD circuit is somewhat different from the usual types: it can operate at the standard frequency of 1000 Hz, but it also is tunable from 970 Hz to 1030 Hz, and has an adjustable *Q* factor of 0.3 to over 50. Op-amp U1, a TLC272 CMOS unit, contains the two voltage-followers required to buffer the input to the bootstrapped twin-T notch filter. Tuning is accomplished by R1, R2, and R3, which are standard linear-taper slide pots "ganged" together by mounting them side-by-side and gluing their sliders together. The only other important construction hint is to use twisted pair at the circuit's input and output.

To calibrate the circuit, input a 1000-V RMS signal at 2000 Hz, set S1 to TEST, and adjust R7 for a reading of 0.99-V RMS on a true-RMS voltmeter at the output.

To use the circuit, set S1 to TUNE, input a 1000-Hz sine-wave signal to the amplifier under test, and set the amplifier's output to the THD adapter and tune R1/R2/R3 for the lowest output signal. Then, set S1 to TEST and read the RMS voltage. To calculate the percent THD use:

$$THD = \left(\frac{V_{\text{out}}}{V_{\text{in}}}\right) \times 100$$

SOUND SUBCARRIER GENERATOR

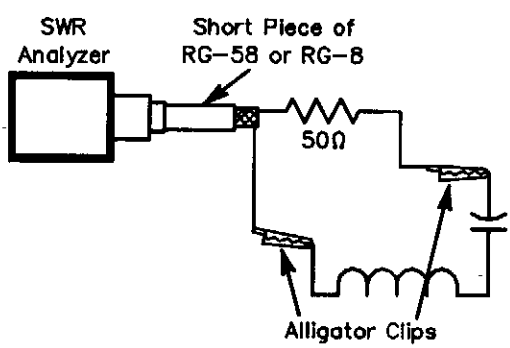


WILLIAM SHEETS

Fig. 50-42

This circuit will generate an FM sound subcarrier at 4.5 or 10.7 MHz for FM and TV IF testing and alignment. Q1 is an audio amplifier and Q2 is a VCO modulated by an MV2112 varactor. Deviation up to 1% of frequency can be obtained. L is chosen to resonate with the circuit capacitance to either 4.5 or 10.7 MHz. The values will be around 2 to 10 μH, depending on the frequency.

INDUCTANCE AND CAPACITANCE DETERMINER WITH SWR BRIDGE



QST

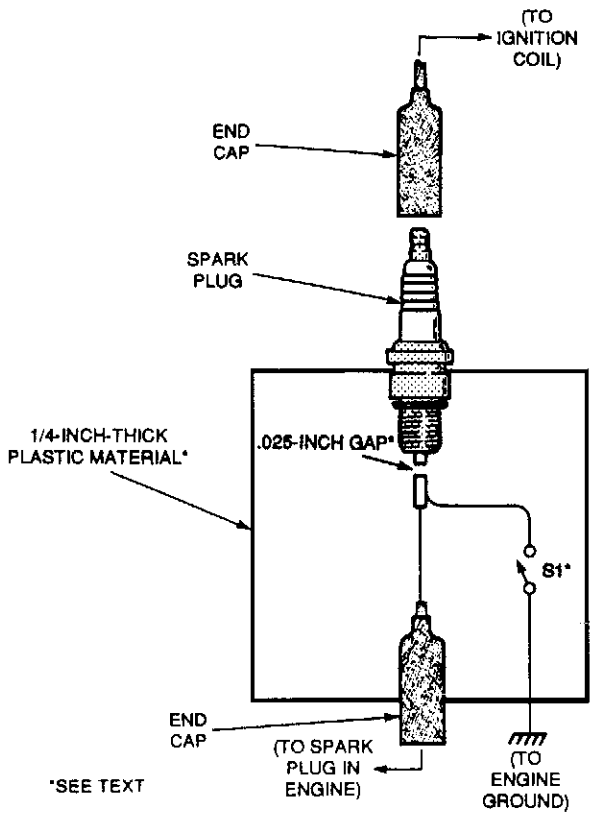
Fig. 50-43

At resonance, the SWR will be 1:1 with a 50-Ω resistance, as reactance is zero. If either L or C is known:

$$|X_c| = |X1| = \frac{1}{2\pi fC} = 2\pi fL \quad (|X_L|)$$

$$L_{\text{unknown}} = \frac{1}{(2\pi f)^2 C} \qquad C_{\text{unknown}} = \frac{1}{(2\pi f)^2 L}$$

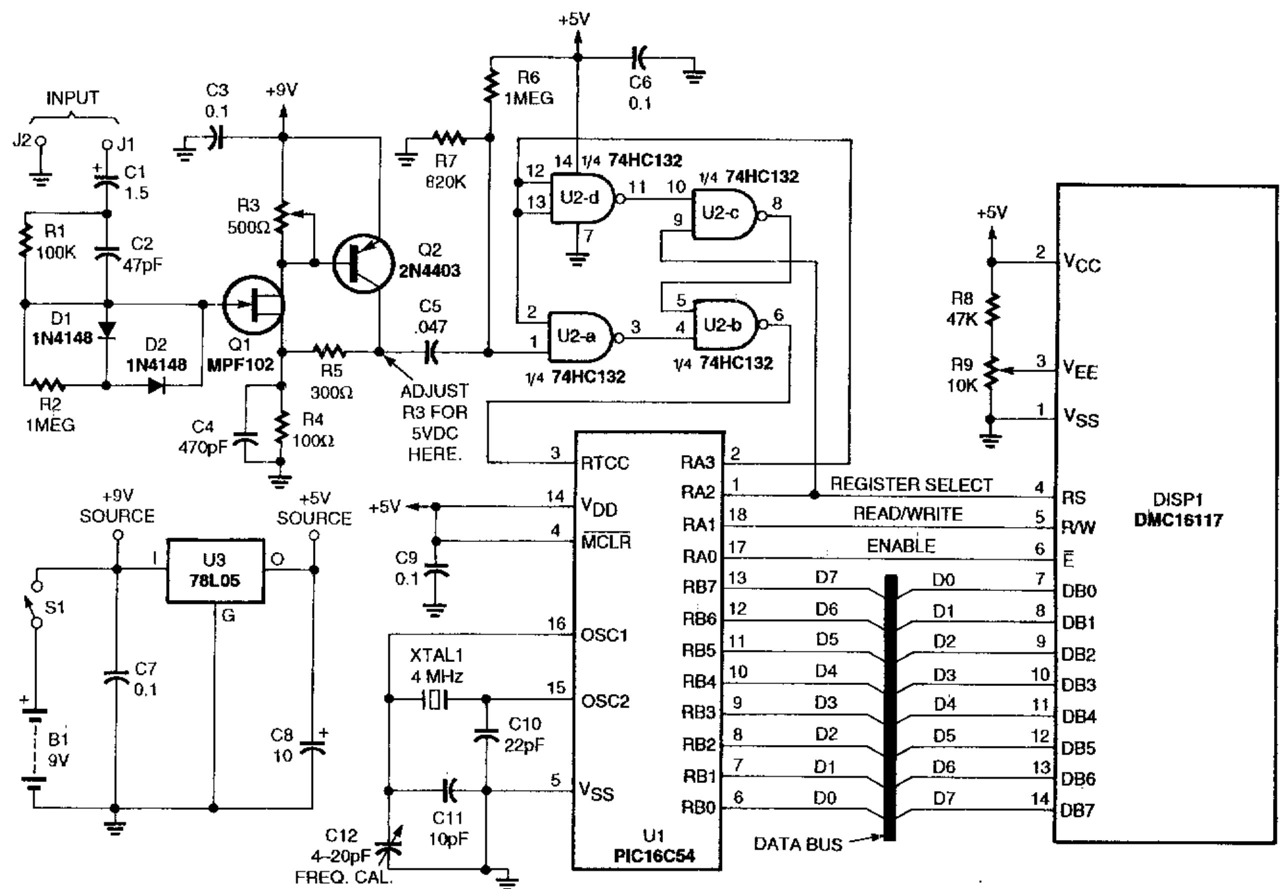
MOTORCYCLE TUNE-UP AID



POPULAR ELECTRONICS *Fig. 50-44*

Performing a tune-up on a newer bike is made a lot easier with this helpful circuit. Because of the high voltages present, make sure that S1 has an insulated handle and that the fixture is grounded. With the ignition turned off, remove one of the spark plug wires and connect it to the spark plug on the fixture. Slip the fixture's end cap over the spark plug on the cycle and you're ready to go. Open S1 and start the engine. Then, close S1; the cylinder with the fixture should not fire and a spark should be seen at the fixed gap. Be sure that the fixture is connected to the engine ground before closing S1.

50-MHz FREQUENCY COUNTER

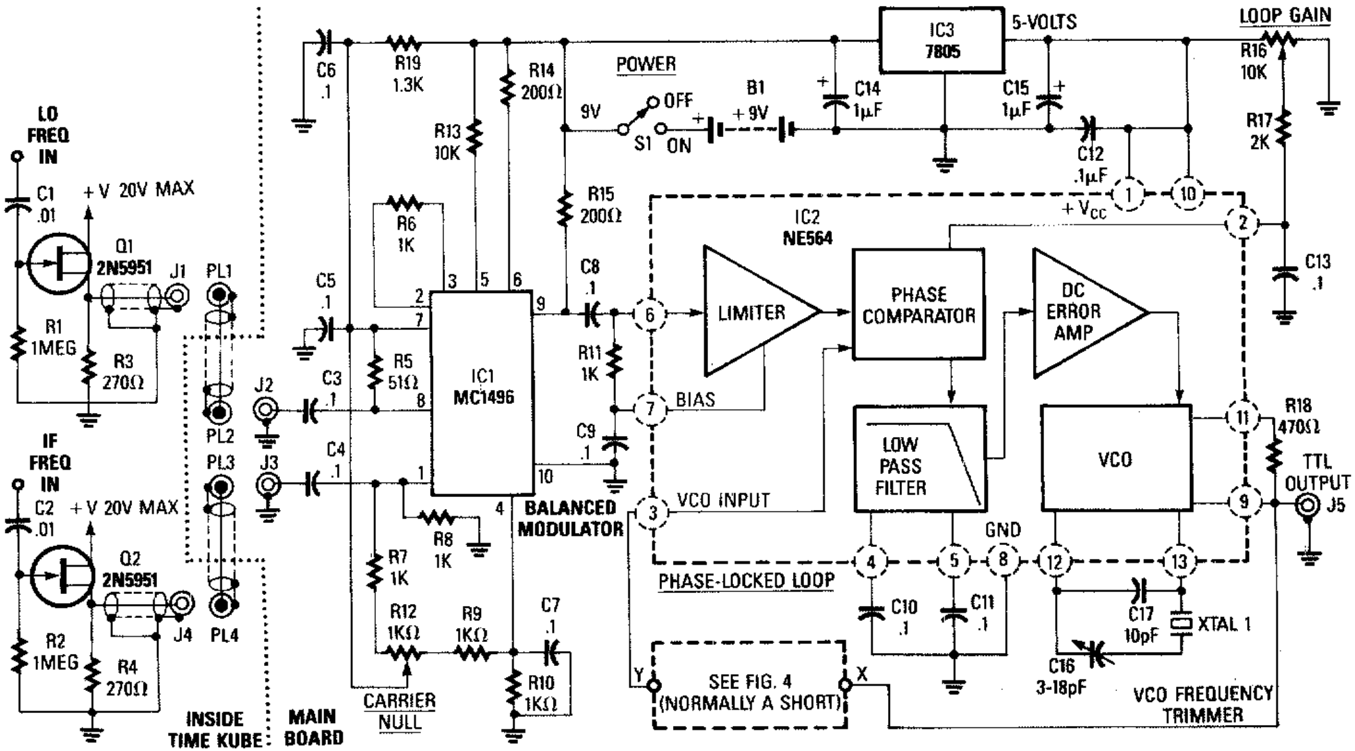


POPULAR ELECTRONICS

Fig. 50-45

This inexpensive frequency counter uses a microcontroller as the counter. The microcontroller feeds an LCD display module that accepts standard ASCII code. The frequency is displayed as Hz, kHz, or MHz and the counter is autoranging.

10-MHz FREQUENCY STANDARD

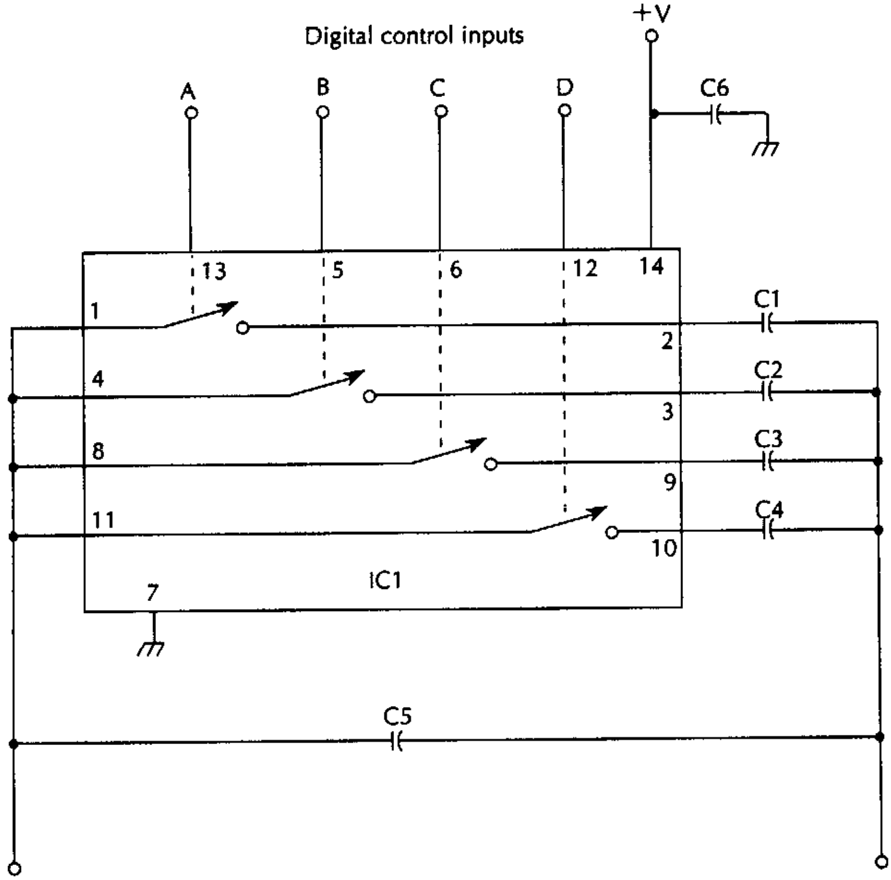


RADIO-ELECTRONICS

Fig. 50-46

A Radio Shack 10-MHz receiver is used as a basis for this circuit. The L.O. and IF frequencies are added. When the receiver is tuned to 10 MHz (WWV), the sum of the L.O. and IF are used to phase lock a VCO to the 10-MHz signal. By using a divider in the loop, 2.5 or 5 MHz can be used as well.

PROGRAMMABLE CAPACITOR CIRCUIT

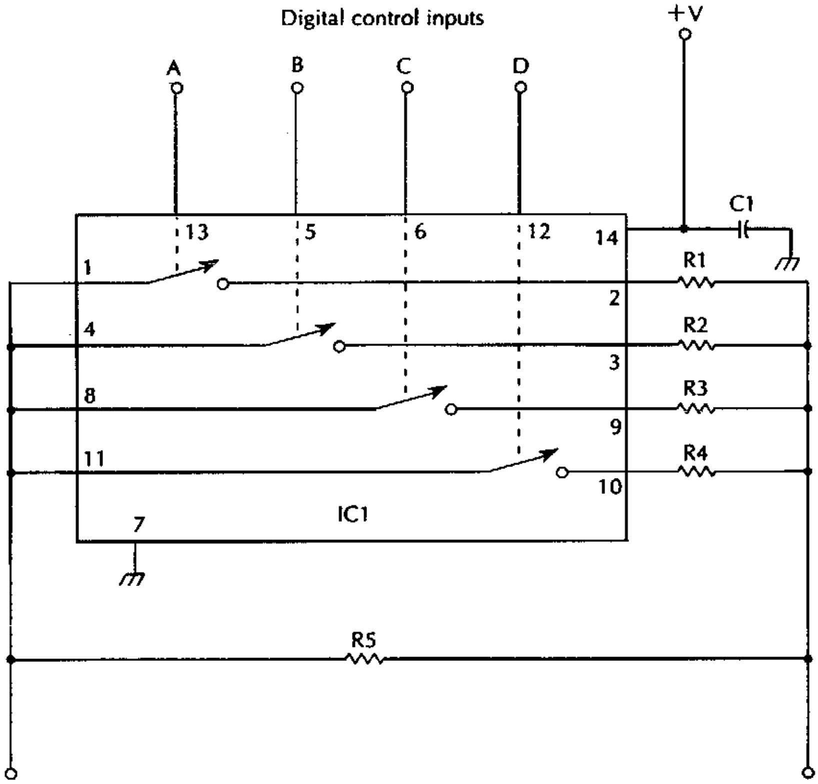


McGRAW-HILL *Fig. 50-47*

IC1	CD4066 quad bilateral switch
C1	10-µF, 25-V capacitor
C2	22-µF, 25-V capacitor
C3	47-µF, 25-V capacitor
C4	100-µF, 25-V capacitor
C5	1-µF, 25-V capacitor
C6	0.1-µF, 25-V capacitor

The programmable capacitor can be very useful in circuits where you need to switch capacitance values. Remember that the "ON" resistance of IC1 appears in series with the capacitors and must be taken into account in some applications as it is not negligible.

PROGRAMMABLE RESISTOR CIRCUIT



MCGRAW-HILL

Fig. 50-48

IC1	CD4066 quad bilateral switch
C1	0.1-µF capacitor
R1	10-kΩ, ¼-W 5% resistor
R2	4.7-kΩ, ¼-W 5% resistor
R3	2.2-kΩ, ¼-W 5% resistor
R4	1-kΩ, ¼-W 5% resistor
R5	1-MΩ, ¼-W 5% resistor

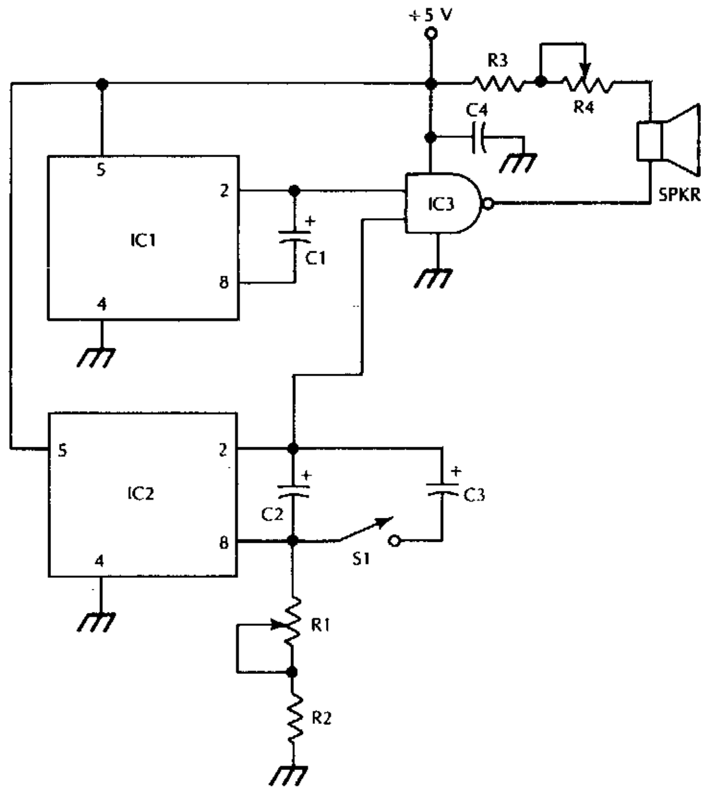
A programmable resistor can replace a potentiometer or fixed resistor. Remember that the "ON" resistance of IC1 might have to be taken into account in some applications.

51

Metronome Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

AUDIBLE METRONOME

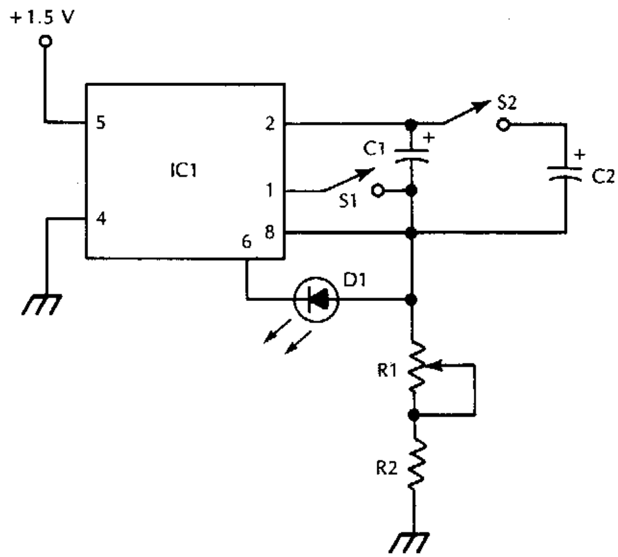


IC1, IC2	LM3909 LED flasher/oscillator	C3	47 μF 10 V electrolytic capacitor
IC3	7400 quad NAND gate	C4	0.01 μF capacitor
SPKR	small loudspeaker	R1	50 kΩ potentiometer
S1	SPST switch	R2	3.3 kΩ 1/4 W 5% resistor
C1	1 μF 10 V electrolytic capacitor	R3	47 kΩ 1/4 W 5% resistor
C2	100 μF 10 V electrolytic capacitor	R4	500 Ω potentiometer

McGRAW-HILL ***Fig. 51-1***

IC1 generates an audible frequency while a variable very low frequency is generated by IC2. R1 sets the metronome rate. The two signals are combined in IC3.

VISUAL METRONOME



IC1	LM3909 LED flasher/oscillator	C2	47 µF 6 V electrolytic capacitor
D1	LED	R1	50 kΩ potentiometer
S1, S2	SPST switch	R2	3.3 kΩ 1/4 W 5% resistor
C1	100 µF 6 V electrolytic capacitor		

McGRAW-HILL

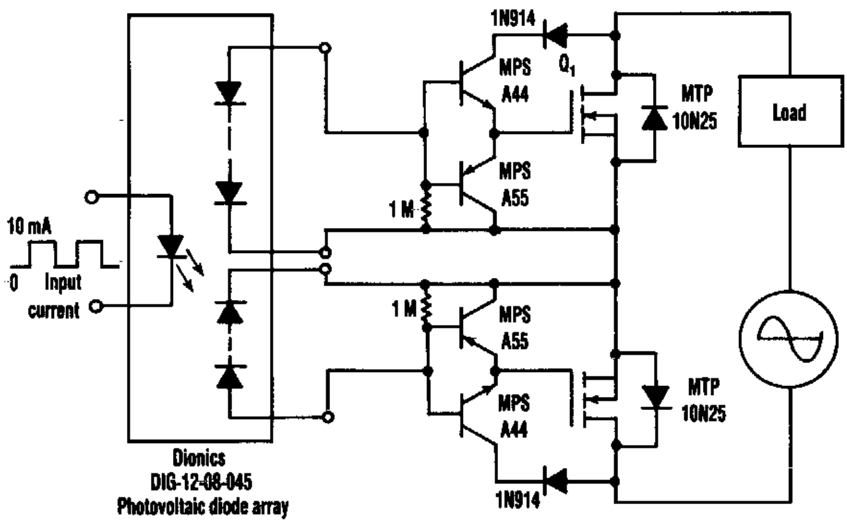
Fig. 51-2

52

Miscellaneous Treasures

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

MOSFET DRIVE CURRENT BOOSTER



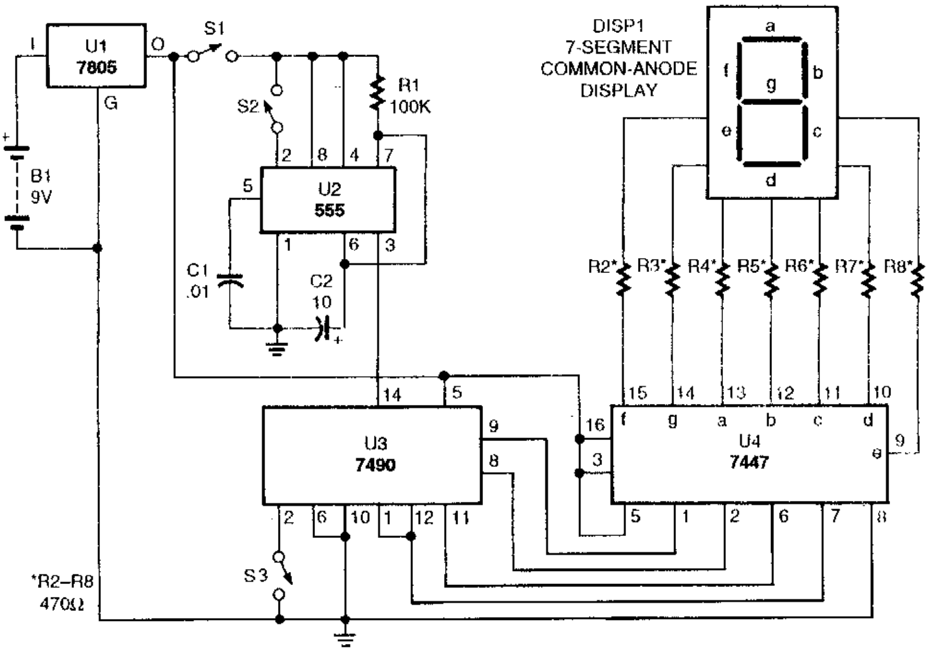
ELECTRONIC DESIGN *Fig. 52-1*

A clean and inexpensive way to eliminate the floating-gate supply is to use the voltage available at the MOSFET's drain to drive its gate. Tying the collector of Q1 (a high-voltage, small-signal, 400-V NPN) to the MOSFET's drain supplies sufficient gate-drive voltage when it's needed most—when the MOSFET's drain-to-source voltage is high. Two such circuits used back-to-back form an ac relay.

Using the emitter follower attached to the drain increases gate-drive current and decreases the MOSFET's turn-on time by a factor equal to the high-voltage NPN's beta. The resulting drain-to-source voltage fall times depend on the MOSFET's size and its required gate charge. The circuit that's used gives a fall time of 200 µs for an MTP10N2f5 10-A, 250-V MOSFET. With such fall times cutting switching losses, pulse-width modulation at frequencies under 100 Hz is possible.

During tune-on, V_{DS} falls rapidly until it reaches the sum of the 1N914 diode's 0.7-V drop, the collector-emitter saturation voltage of Q1, and the gate-to-source voltage required to support the load current. At that point, the diode array completes the MOSFET's turn-on, unaided by the buffer. This slows the fall of V_{DS} considerably when it reaches about 5 to 7 V. In high-voltage, low-frequency systems, tailing of V_{DS} is tolerable because the tail's voltage magnitude constitutes a small fraction of the switching voltage. The IN914 makes it possible for V_{GS} to exceed V_{DS} as the MOSFET completes turn-on.

SIMPLE EVENT COUNTER

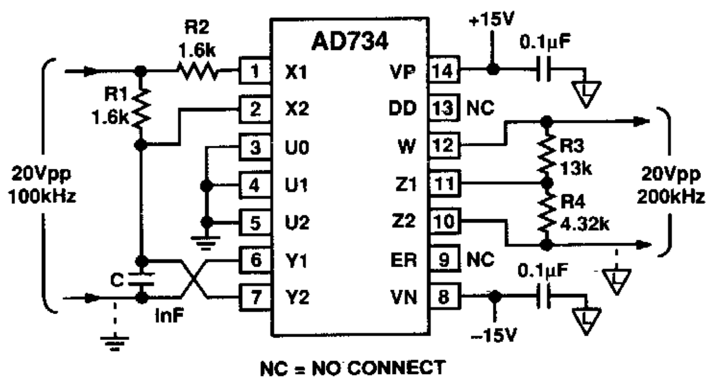


POPULAR ELECTRONICS

Fig. 52-2

S1 is a power switch. U2 drives counter U3 by producing a pulse when S2 is depressed. U4 and DISP1 read the count of counter IC U3. S3 is a reset to zero switch. The counter is a basic one-digit circuit useful as a holding block or by itself.

FREQUENCY DOUBLER

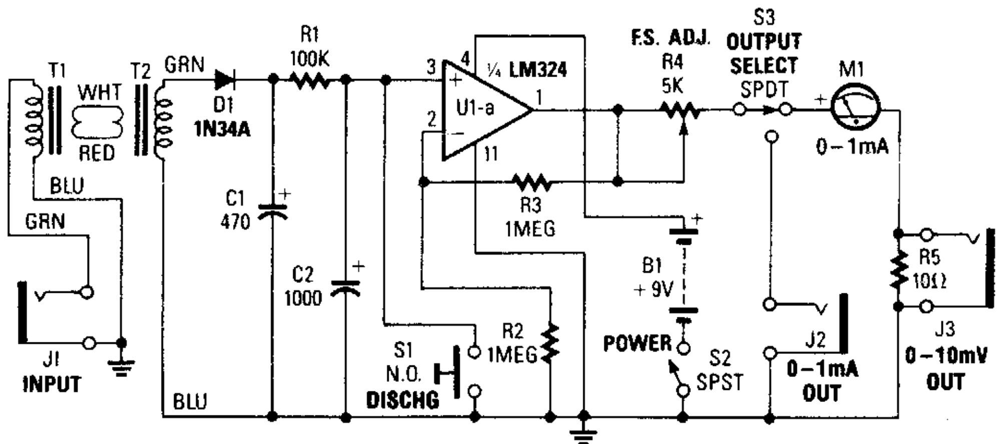


ANALOG DEVICES

Fig. 52-3

An Analog Devices AD734 four-quadrant analog multiplier is used as a frequency doubler.

ATMOSPHERE NOISE MONITOR



POPULAR ELECTRONICS *Fig. 52-4*

Tune an unmodified transistor radio to an unused frequency near 540 kHz that's free of broadcast-station interference; the receiver is used to pick up sferics. The received signal is fed from the receiver's earphone jack through a patch cord to the input jack (J1) of the circuit. The back-to-back audio transformers, T1 and T2, provide a suitable impedance match and signal level when the unit is used with various receivers.

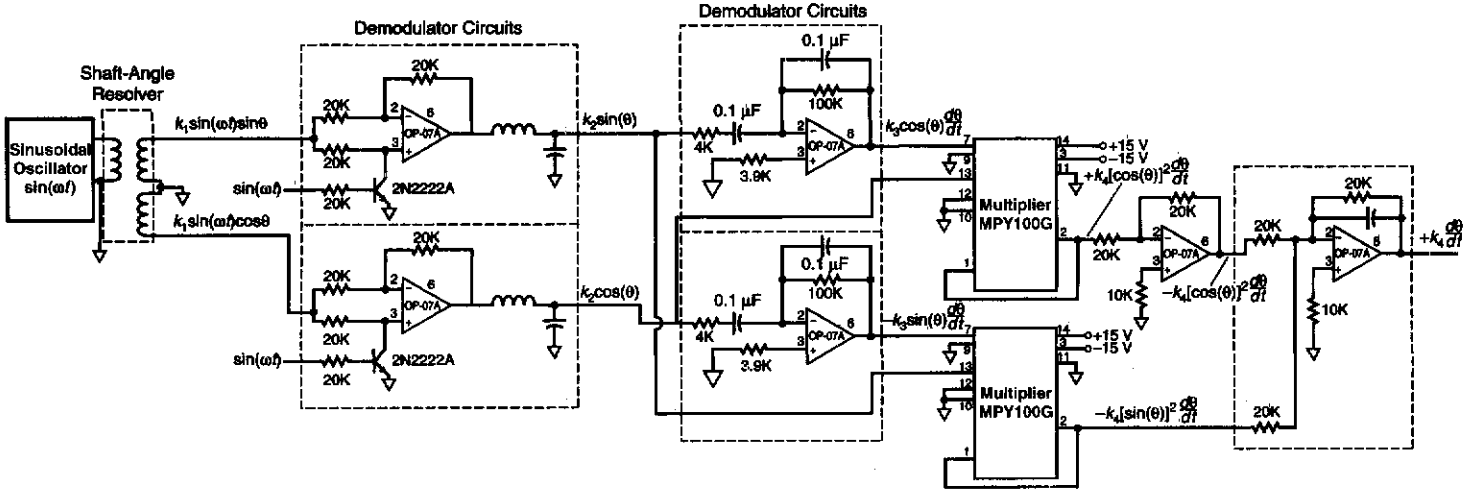
Diode D1 rectifies the audio input from the receiver to pulsating dc, which is filtered by C1, R1, and C2 to provide a time constant of several minutes. That dampens out fluctuations in most cases, unless lightning flashes are very infrequent.

The voltage appearing at the output of the filter is a function of signal strength transferred by C1. Switch S1 is included to provide a convenient way to discharge the capacitors, if adjustments are required during a monitoring session.

Integrated circuit U1 (one section of an LM324 quad pro op amp) is used as a high input-resistance voltmeter. Resistors R2 and R3 determine amplifier gain, and potentiometer R4 is used to adjust full-scale meter deflection for a suitable voltage level at the input. A value of 1.5 V has been satisfactory for use with several receivers tried, but they can be changed.

If the monitor is to be used only as a meter, the milliammeter can be connected directly between R4 and chassis ground, omitting R5, S3, J2, and J3. The latter components provide suitable output for use with a chart recorder having a full-scale range of either 10 mV or 1 mA. The circuit, when powered from a 9-V battery, draws about 1 mA.

TACHOMETER DERIVED FROM BRUSHLESS SHAFT ANGLE RESOLVER



NASA TECH BRIEFS

Fig. 52-5

The tachometer circuit operates in conjunction with a brushless shaft-angle resolver. By performing a sequence of straightforward mathematical operations on the resolver signals and utilizing a simple trigonometric identity, it generates a voltage proportional to the rate of rotation of the shaft.

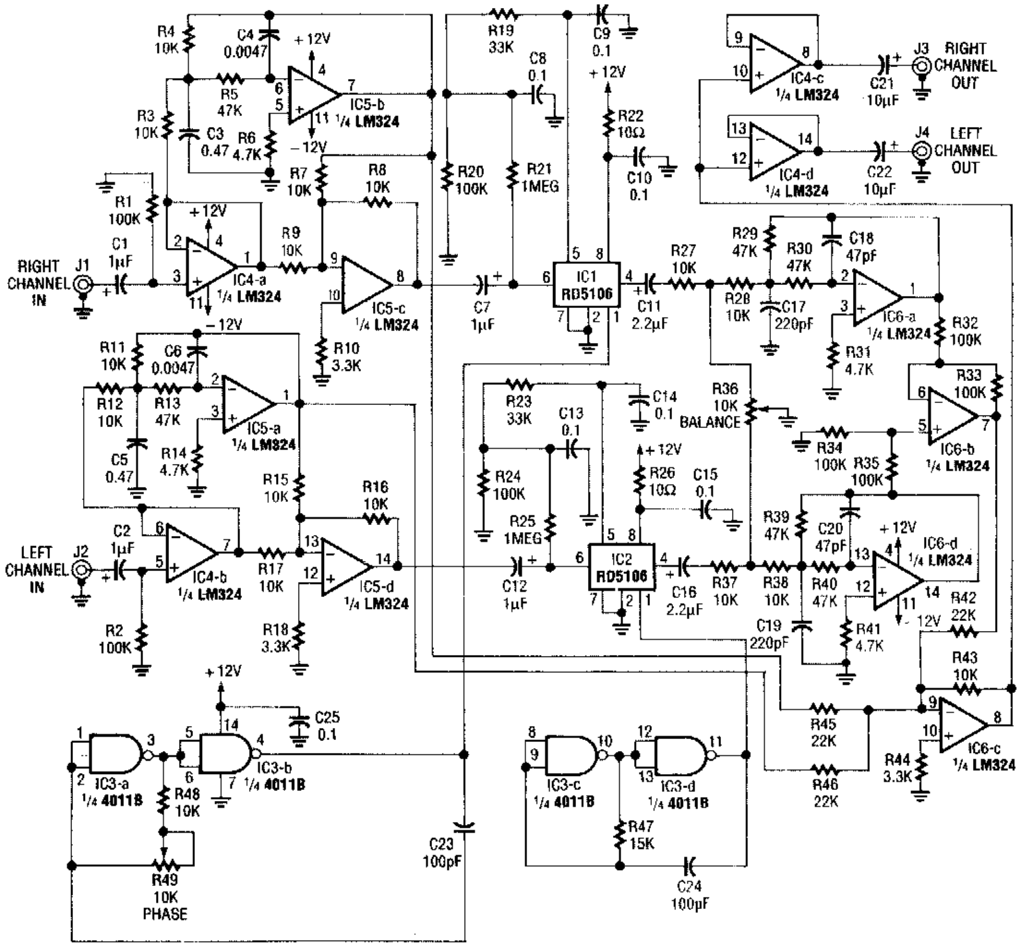
The figure illustrates an analog tachometer circuit that processes the input and output signals of a two-phase, brushless, transformer-type shaft-angle resolver into a signal with instantaneous amplitude proportional to the instantaneous rate of rotation of the shaft. The processing in this circuit effects a straightforward combination of mathematical operations leading to a final operation based on the well-known trigonometric identity $[\sin(x)]^2 + [\cos(x)]^2 = 1$ for any value of x.

The resolver is excited with a periodic waveform; a sinusoid is indicated in the figure, but a square, triangular, or other periodic waveform could be used instead. Thus, the two outputs of the resolver are $k_1 \sin(\omega t)\sin(\theta)$ and $k_1 \sin(\omega t)\cos(\theta)$, where k_1 is a constant proportional to the amplitude of excitation, ωt is $2\pi x$ the frequency of excitation, t is time, and θ is the instantaneous shaft angle.

The two outputs of the resolver are then processed, along with a replica of the sinusoidal excitation, by demodulators. These signals are then differentiated with respect to time in two differentiator circuits. Notice that $d\theta/dt$ is the rate of change of the shaft angle and is the quantity that one seeks to measure.

Next, a multiplier circuit forms a product of the demodulator and differentiator outputs proportional to $\sin(\theta)$, and the product of the demodulator and differentiator outputs proportional to $\cos(\theta)$. The output of the cosine multiplier is fed to a unit-gain inverting amplifier.

VOCAL STRIPPER

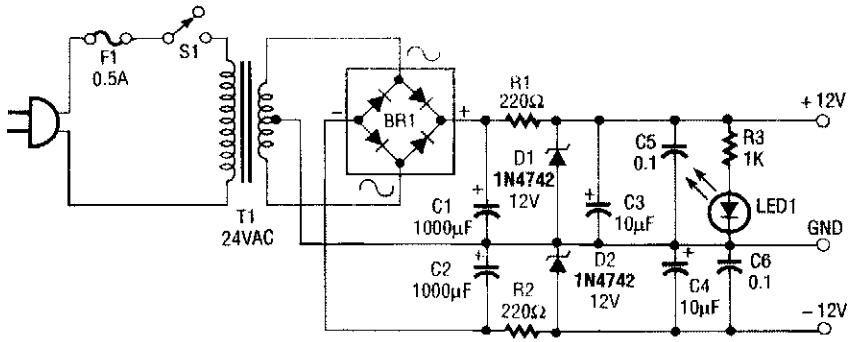


RADIO-ELECTRONICS *Fig. 52-6*

The schematic of the lead vocal filter is shown in the figure. The left and right channel signals are coupled through C1 and C2 to buffer amps IC4-a and IC4-b. From the buffer amps, the left and right channel signals pass through active crossovers IC5-a and IC5-b, sending all low frequencies to a final mixer IC6-c, and all middle and high frequencies to analog delay lines IC1 and IC2, RD5106 256-sample bucket-brigades. Integrated circuit IC2 delays the left-channel signal by 2.4 ms, set by the fixed-frequency clock generated by ½IC3, R47, and C24. The right channel signal is delayed by IC1 with a variable-frequency clock generated by ½IC3, R48, R49, and C23. Potentiometer R49 is used for phase adjustment.

The output of each delay line from IC1 and IC2 passes through low-pass filters IC6-a and -d, and their associated parts, to filter out high-frequency sample-steps produced by IC1 and IC2. Balance control R36 is adjusted for equal amplitude of the left and right channels. IC6-b is a difference amplifier that cancels all lead vocals that are common to both channels. The resulting signal from IC6-b is remixed with low frequencies by IC6-c and is then sent to the output via buffers IC4-c and IC4-d.

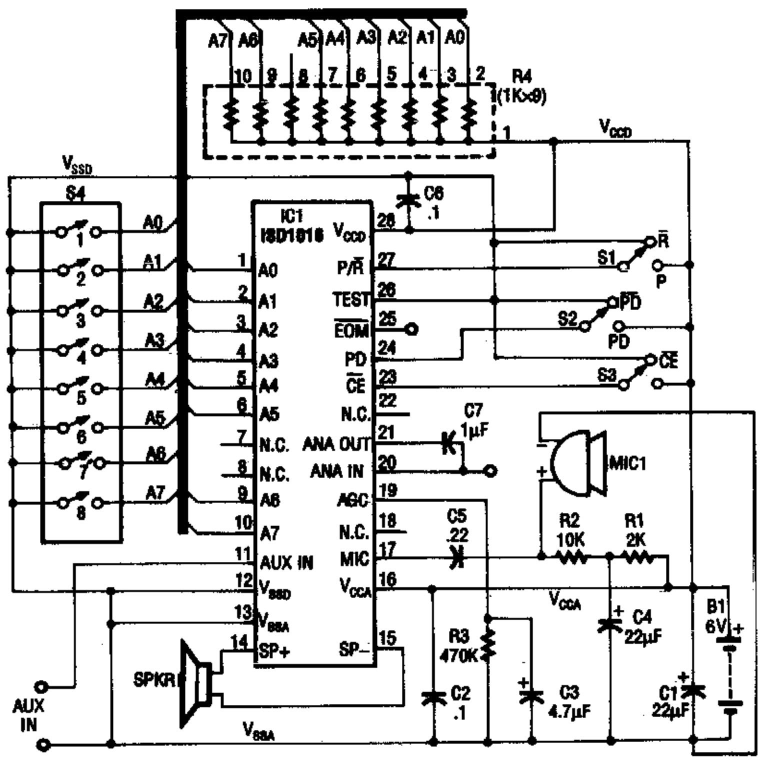
VOCAL STRIPPER POWER SUPPLY



RADIO-ELECTRONICS *Fig. 52-7*

The power supply schematic for the lead vocal filter circuit.

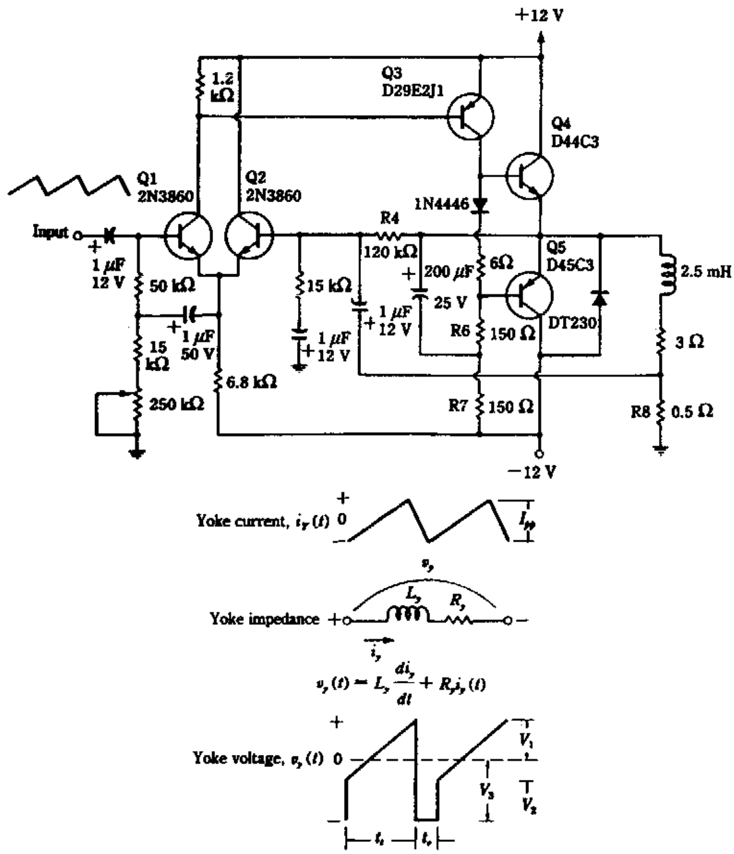
SINGLE-CHIP MESSAGE SYSTEM



ELECTRONICS EXPERIMENTERS HANDBOOK *Fig. 52-8*

The ISD1016 is a complete analog audio record/playback system on a chip. The analog signal is sampled and the samples stored in an EEPROM as analog levels. Upon playback, the analog data is read out and amplified. Up to 16 seconds of data (audio) can be stored.

TELEVISION VERTICAL DEFLECTION CIRCUIT



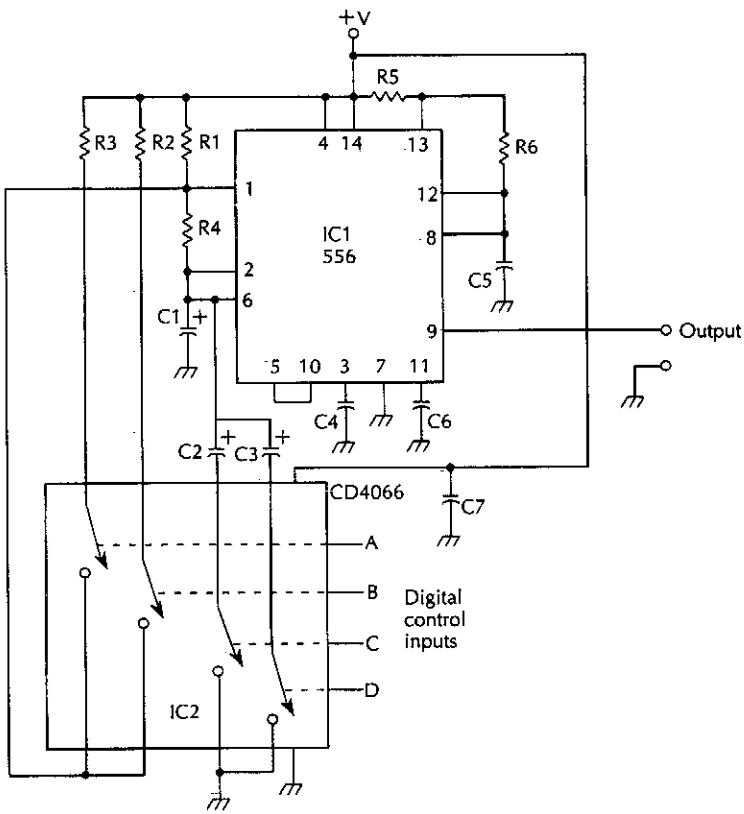
$$v_y(t) = L_y \frac{di_y}{dt} + R_y i_y(t)$$

MCGRAW-HILL

Fig. 52-9

Two transistors are used to drive the yoke (2.5 mH + 0.3 Ω) in this deflection circuit. R8 samples the yoke current and provides feedback to Q2, resulting in a very linear current ramp through the yoke.

TONE BURST GENERATOR



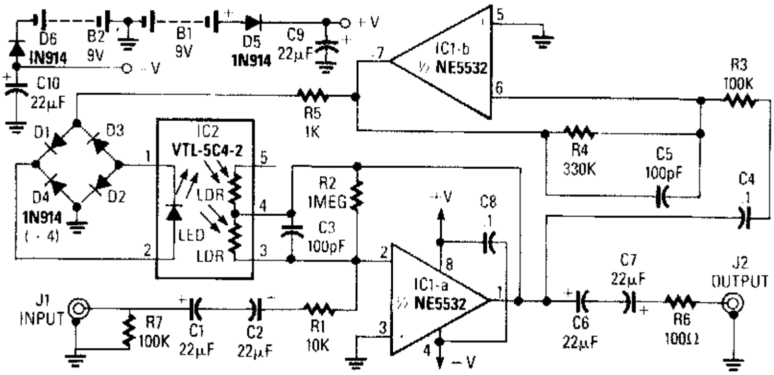
McGRAW-HILL

Fig. 52-10

The burst length is digitally controlled by inputs A, B, C, D. This input selects the necessary values of *R* and *C*. The circuit shown in the figure generates a burst of tone followed by a silent period, then another tone burst is sounded and so forth. The timing of the tone bursts is digitally controlled via the CD4066 (IC2). A parts list for this circuit is given in the table.

IC1	556 dual timer (or two 555 timers)	C5	0.047-µF capacitor
IC2	CD4066 quad bilateral switch	R1	100-kΩ, ¼-W 5% resistor
C1	1-µF, 25-V electrolytic capacitor	R2, R4	220-kΩ, ¼-W 5% resistor
C2	4.7-µF, 25-V electrolytic capacitor	R3	680-kΩ, ¼-W 5% resistor
C3	10-µF, 25-V electrolytic capacitor	R5	12-kΩ, ¼-W 5% resistor
C4, C6	0.01-µF capacitor	R6	4.7-kΩ, ¼-W 5% resistor

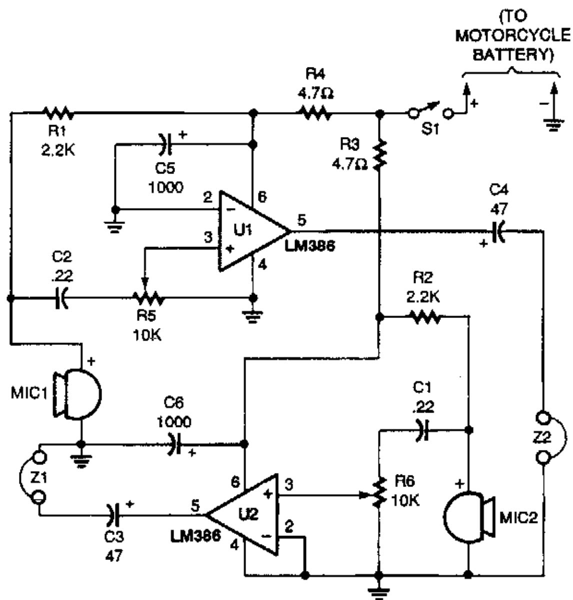
AUDIO VOLUME LIMITER



RADIO-ELECTRONICS ***Fig. 52-11***

In this circuit, amplifier IC1-a provides signal amplification of –40 to +40 dB depending on the valuc of thc LDR. The LDR (light dependent resistor) is driven by rectified audio from voltage follower IC1b and bridge rectifier D1 through D4.

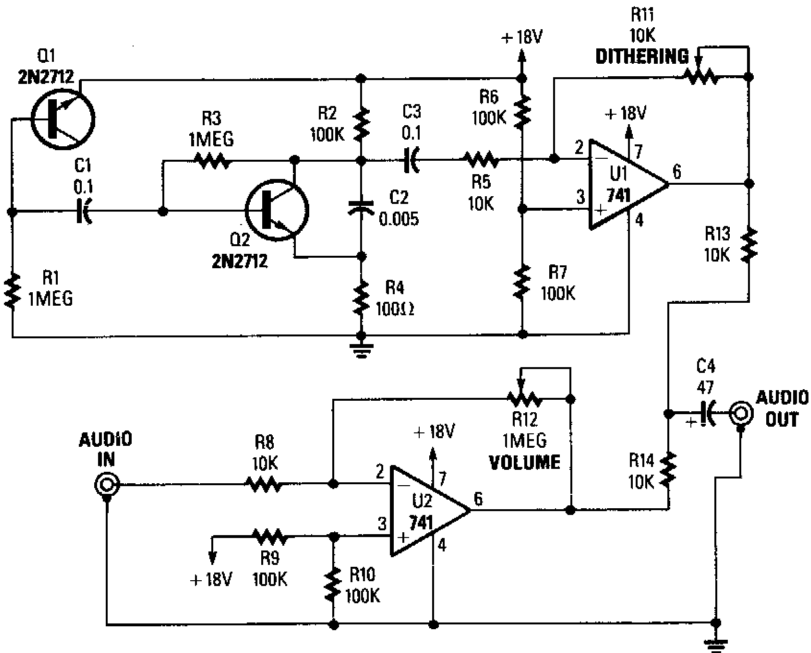
SIMPLE INTERCOM FOR NOISY ENVIRONMENTS



POPULAR ELECTRONICS ***Fig. 52-12***

This intercom was originally designed for motorcycle to passenger communications. A simple "passenger-to-pilot" intercom circuit is shown. Two LM386 ICs are connected in a low-gain amplifier circuit with the headphone output of one paired to the microphone input of the other. The microphones are electret elements and the earphones can be of the in-ear type or of the small stereo/mono type that will fit inside a helmet. Both amplifiers in the circuit operate at a minimum gain of 20 dB.

DITHERIZER

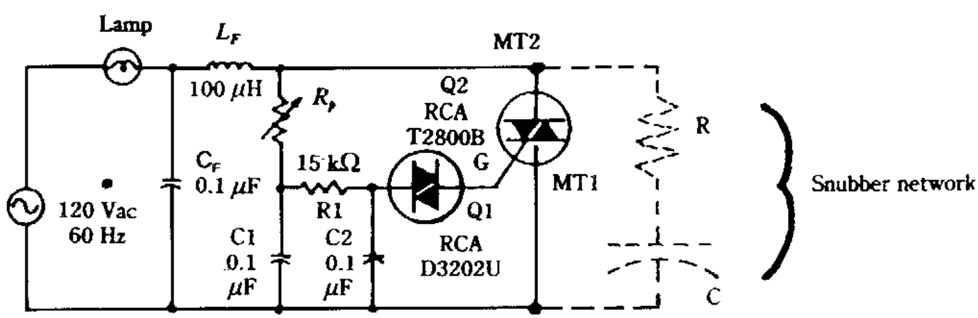


POPULAR ELECTRONICS

Fig. 52-13

In digital audio, a noise signal of amplitude less than one significant bit is often added to the audio to reduce the quantizing effect and improve the audio quality by trading digital "noise" for analog noise, which does not have the harsh sound. This circuit consists of a noise generator to add a low level of noise to an analog signal to be digitized, or an analog signal from a digital source.

TRIAC LAMP DIMMER CIRCUIT



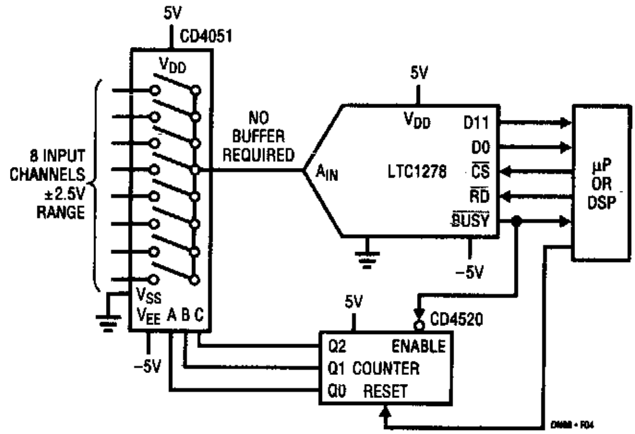
For 220-V, 50/60-Hz Operation, replace T2800B with T2800D.

McGRAW-HILL

Fig. 52-14

The brightness of a lamp or lamps can be varied with this circuit. The snubber circuit values are typically 0.1 μF and 100-Ω. R_8 is typically 25 to 100 kΩ.

500-ksps 8-CHANNEL DATA ACQUISITION CIRCUIT

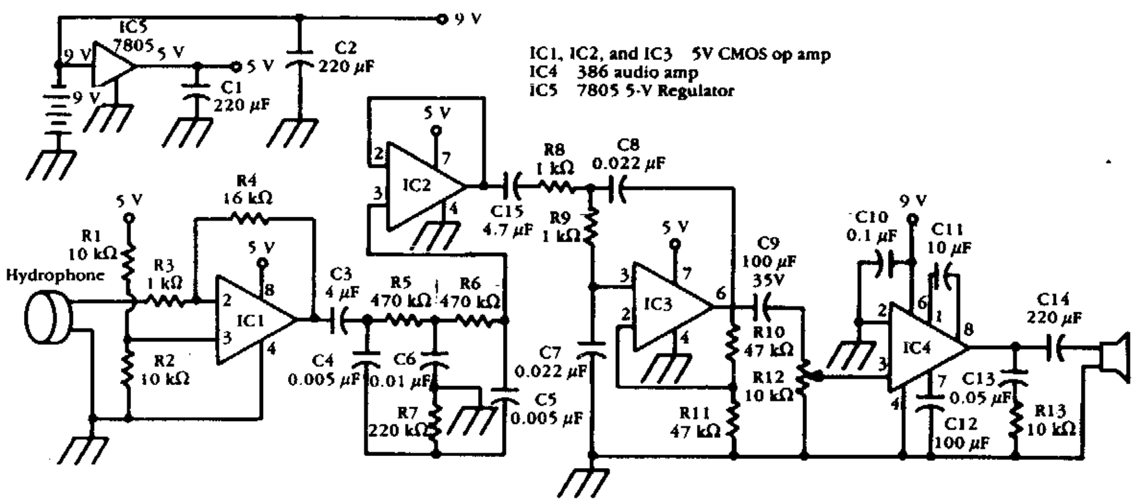


LINEAR TECHNOLOGY

Fig. 52-15

The high input impedance of the LTC1278 allows multiplexing without a buffer amplifier. Both single channel and multiplexed high-speed data acquisition systems benefit from the LTC1278/LTC1279's dynamic conversion performance. The 1.6-µs and 1.4-µs conversion and 200-ns and 180-ns S/H acquisition times enable the LTC1278/LTC1279 to convert a 500 ksps and 600 ksps, respectively. The figure shows a 500-ksps 8-channel data acquisition system. The LTC1278's high input impedance eliminates the need for a buffer amplifier between the multiplexer's output and the Adc's input.

HYDROPHONE

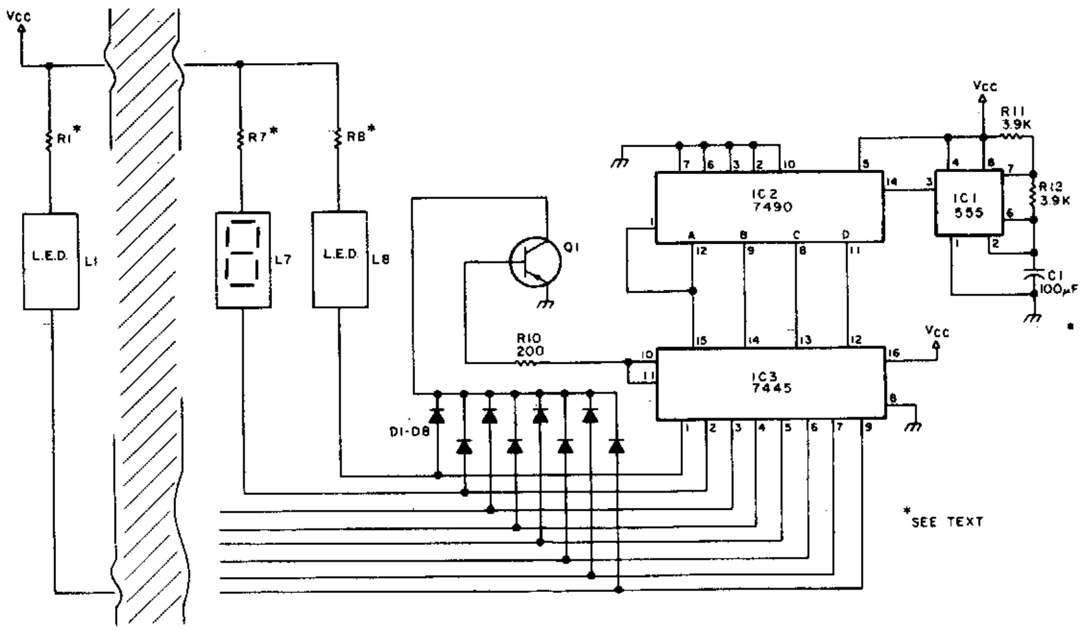


McGRAW-HILL

Fig. 52-16

A commercially available hydrophone transducer is used in this system. The transducer is connected via a cable to the amplifier, which remains out of the water. The hydrophone should be suitably mounted for intended application.

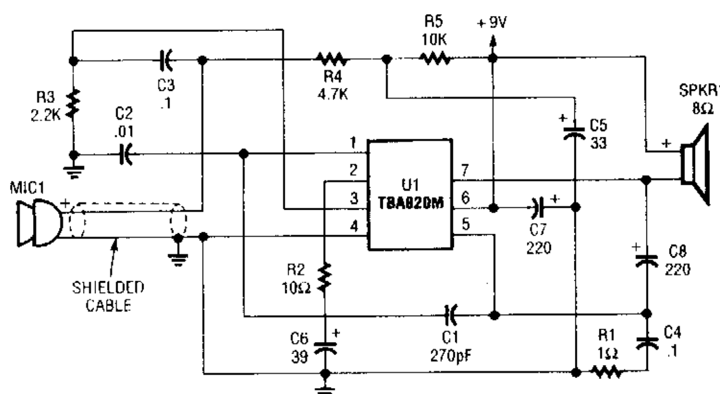
YOUR NAME IN LIGHTS



73 AMATEUR RADIO TODAY ***Fig. 52-17***

This circuit will enable you to put a name or callsign in lights using seven-segment LEDs. The display will spell the desired name out sequentially. Select the correct type of LED. Solder the correct leads together to form the letters you want. After mounting the appropriate current-limiting resistor, the 7445 can only sink 80 mA, so a PNP transistor is needed to handle the current required to light the letters. The heart of the circuit is a 555 oscillator into a 7490 decade counter, which is decoded by a 7445 open-collector driver chip.

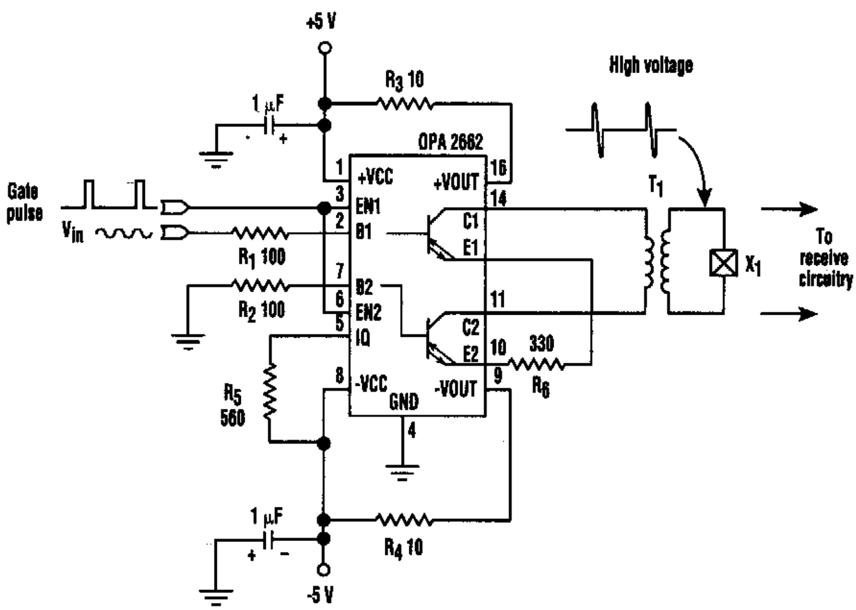
UNDERWATER MICROPHONE



ELECTRONIC HOBBYISTS HANDBOOK ***Fig. 52-18***

This circuit uses a TBA820 audio IC to amplify underwater sounds. The microphone must be waterproofed. This project was originally used in a home aquarium to monitor fish sounds.

PULSE ECHO DRIVER



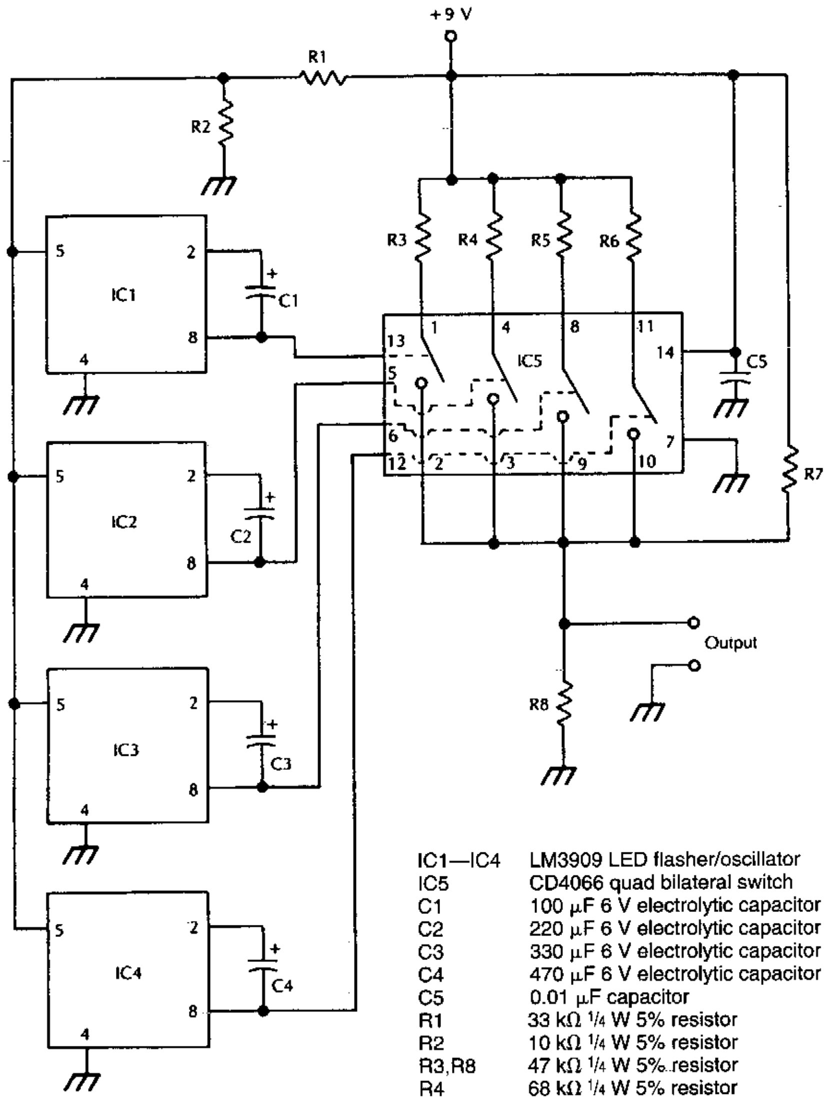
ELECTRONIC DESIGN *Fig. 52-19*

This pulse-echo driver uses the OPA2662 dual operational transconductance amplifier (OTA) from Burr-Brown (the receive circuitry isn't shown). The OTA is preferable over an op amp for driving low impedances because it provides a current output rather than a voltage output.

Ultrasonic pulse-echo applications often incorporate a transformer-coupled crystal to obtain a high-voltage pulse because the echo can be orders of magnitude smaller in amplitude. The transformer turns ratio also provides tuning at the resonant frequency of the crystal, which usually means a relatively low-impedance primary winding.

An operational transconductance amplifier (OTA) is preferred over an op amp to drive such a low impedance. One particular application involves a pulse-echo driver circuit using the OPA2662.

SIMPLE PSEUDORANDOM VOLTAGE SOURCE

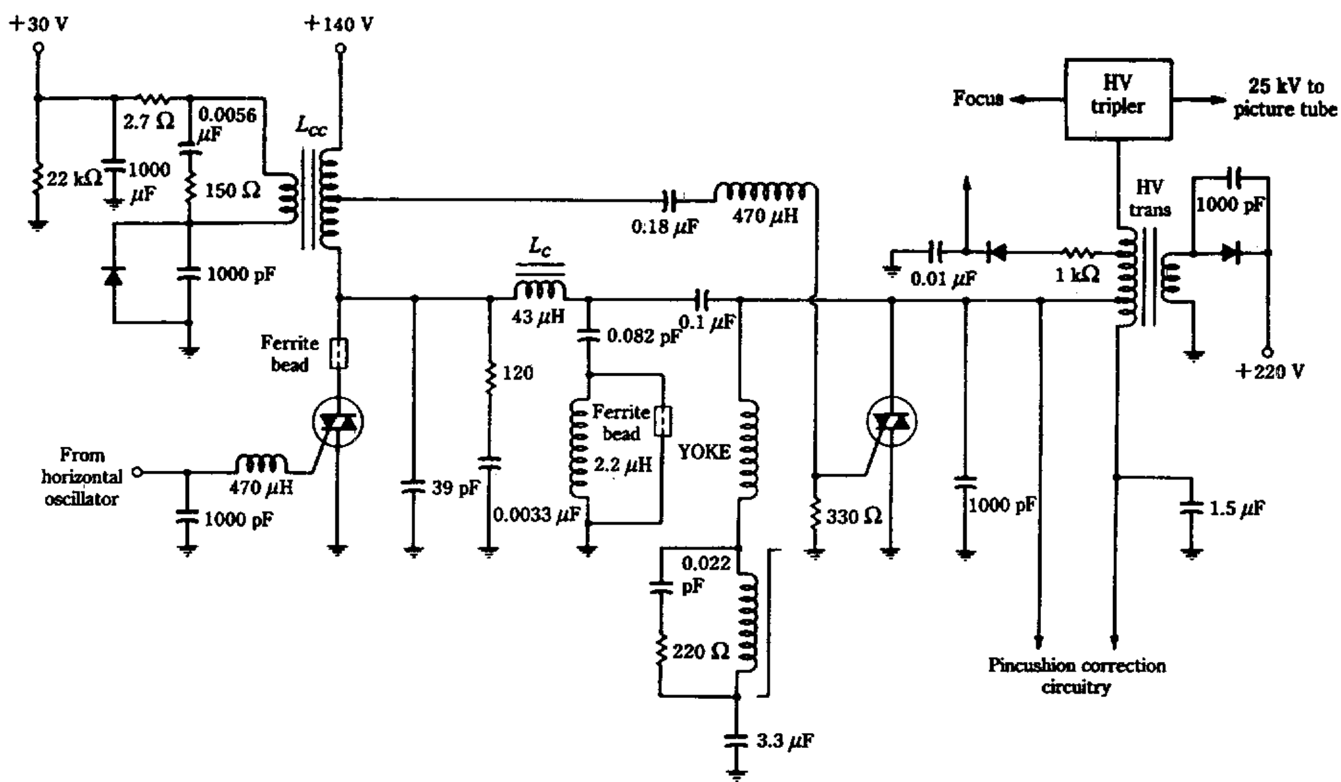


IC1—IC4	LM3909 LED flasher/oscillator
IC5	CD4066 quad bilateral switch
C1	100 μF 6 V electrolytic capacitor
C2	220 μF 6 V electrolytic capacitor
C3	330 μF 6 V electrolytic capacitor
C4	470 μF 6 V electrolytic capacitor
C5	0.01 μF capacitor
R1	33 kΩ 1/4 W 5% resistor
R2	10 kΩ 1/4 W 5% resistor
R3,R8	47 kΩ 1/4 W 5% resistor
R4	68 kΩ 1/4 W 5% resistor
R5	100 kΩ 1/4 W 5% resistor
R6	220 kΩ 1/4 W 5% resistor
R7	1 MΩ 1/4 W 5% resistor

McGRAW-HILL *Fig. 52-20*

An approximation to a pseudorandom voltage is produced by combining the outputs of four low-frequency oscillators with 0.3, 0.6, 0.9, and 1.4 Hz frequencies. The summing network is a quad bilateral switch and resistor network.

TV HORIZONTAL DEFLECTION CIRCUIT

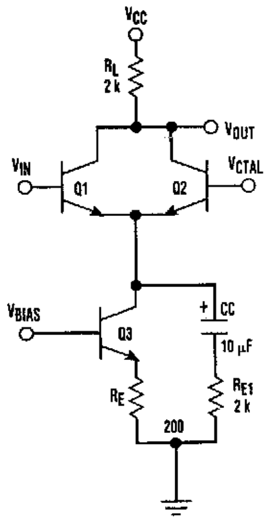


McGRAW-HILL

Fig. 52-21

The circuit illustrates the method of using two SCR devices in a TV horizontal deflection application. This circuit was widely used by certain TV manufacturers as an alternate to the vacuum tube or transistor deflection circuit.

MUTING CIRCUIT



Notes :

Q1, Q2 should be matched ;

Q3 is not critical .

V_{CTRL} low, $\frac{V_{OUT}}{V_{IN}} \approx -\frac{R_L}{R_{E1}}$.

V_{CTRL} high, $\frac{V_{OUT}}{V_{IN}} \to 0$.

$V_{OUTDC} \to$ constant .

ELECTRONIC DESIGN

Fig. 52-22

The circuit operates as follows: The signal is input to V_{in} and a dc control voltage is applied to V_c. V_{bias} determines the desired bias point current. Assuming the following component and voltage values:

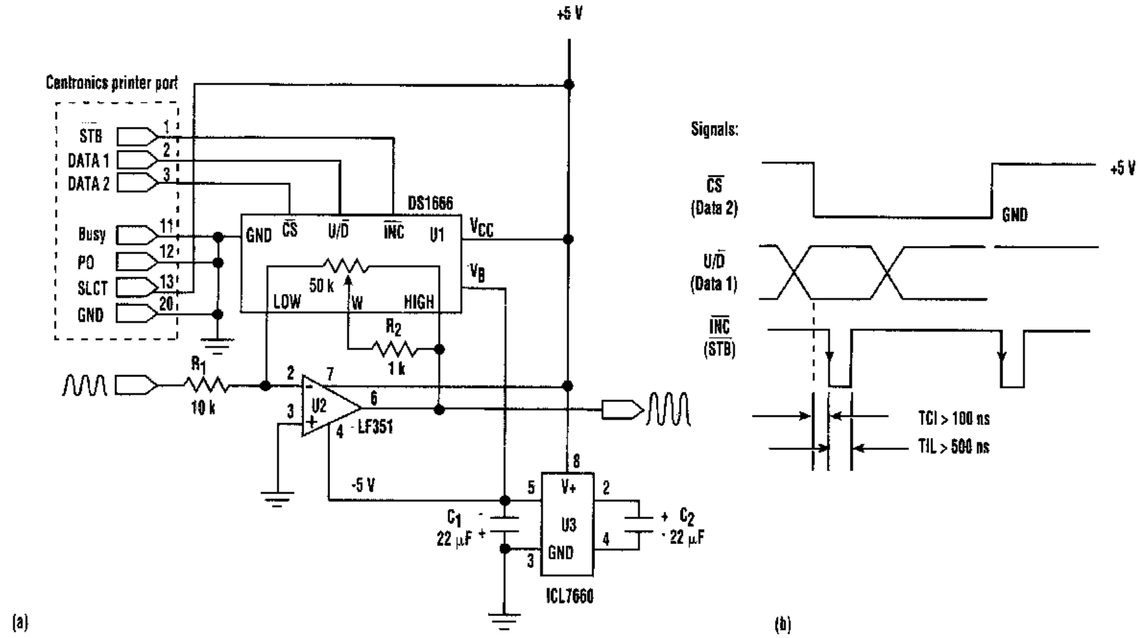
V_{cc} = 7.6 Vdc
V_{bias} = 1 Vdc
V_{in} = 1 Vp-p, ac signal centered about 3.8 Vdc bias
R_L = 2 kΩ
R_E = 200 kΩ
R_{E1} = 2 kΩ

Q3 bias current is 1 mA, and dc output voltage is about 5.8 V with an ac gain of about –1, Q1 and Q2 form a current switch and Q3 acts as a constant current source.

For unmuted operation, V_c = 0 Vdc, and all of the bias current flows through Q1. Consequently, the circuit operates as a normal common emitter stage, with ac gain = $-R_L/R_{E1}$. When V_c = 5 Vdc, all of the bias current flows through Q2, reducing the signal gain to zero. However, because the same dc current flows through R_L in both cases (unmuted and muted), the bias point at the output remains fixed. The C_c/R_{E1} network is required to bypass the Q3 current source (which is a high impedance) to achieve a low ac impedance at the emitter of the Q1 common emitter stage during unmuted operation. C_c is chosen to be a short circuit at signal frequencies of interest. The circuit works best if the Q1 and Q2 pair is matched. Typical change in the output dc voltage from unmuted to muted condition is <5 mVdc.

R_L, R_E, and V_{bias} are chosen for desired dc operating conditions and signal dynamic range. V_{bias} can be generated via a V_{cc} voltage divider. The signal at V_{in} can be ac coupled, but a bias circuit must be added to Q1's base to generate a dc component. R_{E1} is chosen for desired ac gain. V_{in} must be centered about a dc component, and, to assure proper switching action, $V_{CT}R_L$ must be higher than V_{in} by an amount greater than one V_{BE} drop.

SIMPLE REMOTE GAIN CONTROL



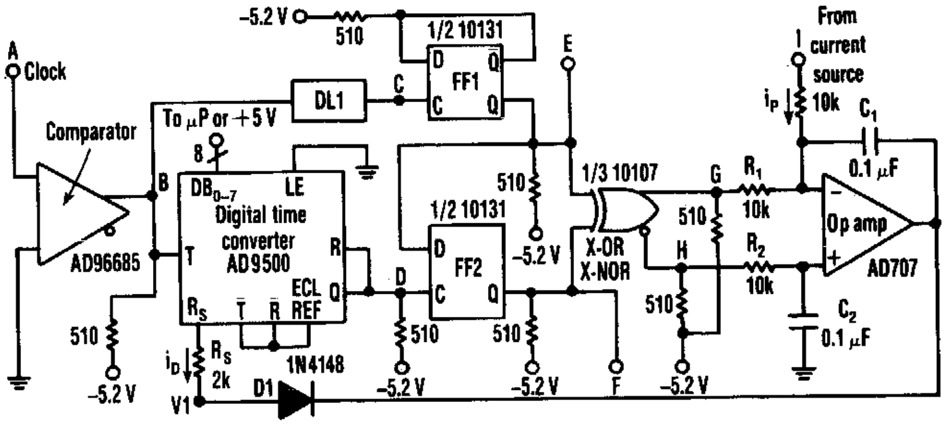
REMOTE CONTROL OF GAIN using a digital audio resistor is possible with this circuit scheme (a). It requires only three ICs and a single 5-V supply to provide gain control via a PC printer port. The input signals to U1 from the printer port are given (b).

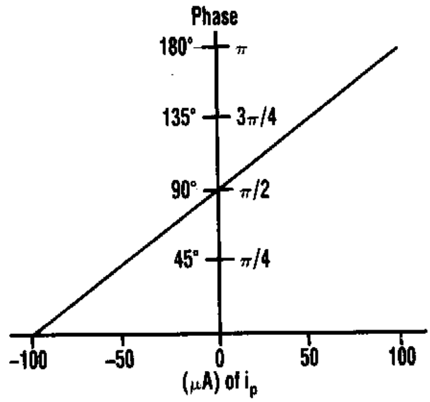
```
100 REM LPGAIN.BAS
120 REM GAIN CONTROL FROM PC PRINTER PORT
200 OPEN "LPT1:" FOR OUTPUT AS #1
210 INPUT "GAIN UP OR DOWN (1/O): "; X
220 IF X<0 OR X>1 THEN GOTO 210
300 REM
310 INPUT "NUMBER OF COUNTS: "; C
320 REM SUBROUTINE WOULD START HERE
330 PRINT #1, CHR$(X);
340 C = C-1 : IF C>0 THEN GOTO 330
350 PRINT #1, CHR$(3)
360 GOTO 210
380 RETURN
```

ELECTRONIC DESIGN *Fig. 52-23*

The listing is a test program that demonstrates circuit operation using an IBM-compatible PC. To form a subroutine for a main program, use lines 330 to 380, deleting line 360. The calling program then would pass values for X (wiper direction) and C (number of increments).

LOOP OSCILLATOR ELIMINATOR





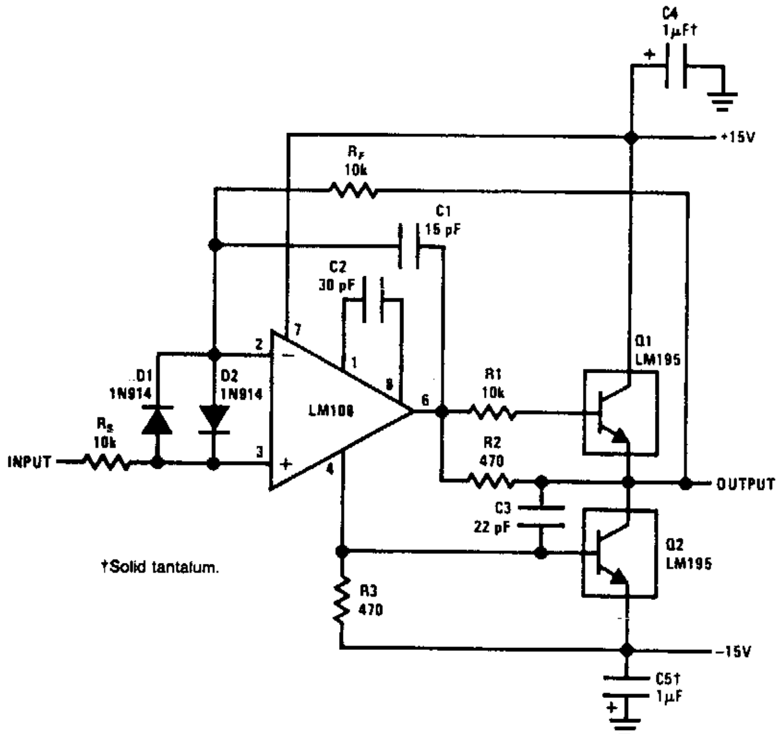
ELECTRONIC DESIGN

Fig. 52-24

This circuit uses negative feedback to a digital-to-time converter, and can supply a current-controlled delay to replace the oscillator in a phase-locked loop that handles input frequencies from 40 kHz to 40 MHz.

A current sourced into the inverting input of the op-amp integrator's summing node can phase shift the pulses at F in relation to those at E by up to 180°.

1-A VOLTAGE FOLLOWER



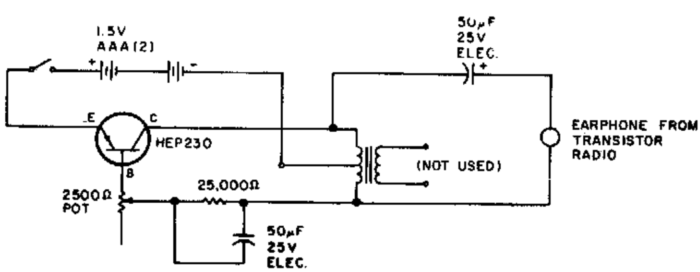
†Solid tantalum.

NATIONAL SEMICONDUCTOR

Fig. 52-25

This power voltage follower is good to 300 kHz.

ELECTRONIC FISH LURE

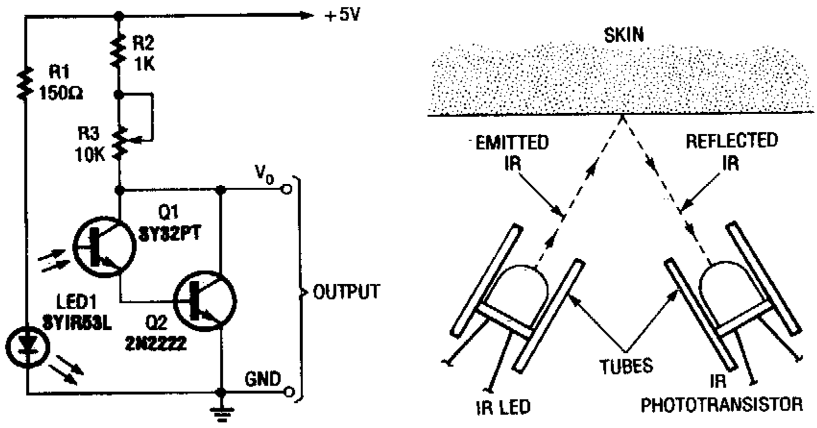


73 AMATEUR RADIO TODAY

Fig. 52-26

The click-click sound lures fish to the vicinity, where your bait or lure can do the rest. The transformer is a subminiature type with a 500-Ω, center-tapped primary and a 3.2-Ω secondary. Put the circuit in a watertight container and lower it into the water.

HEARTBEAT TRANSDUCER

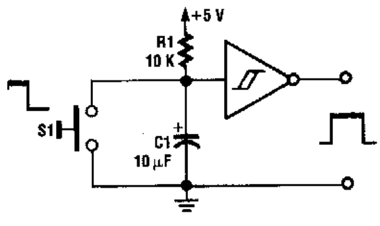


POPULAR ELECTRONICS

Fig. 52-27

A simple heart-beat transducer can be made from an infrared LED and an infrared phototransistor. It works because skin acts as a reflective surface for infrared light. The IR reflectivity of one's skin depends on the density of blood in it. Blood density rises and falls with the pumping action of the heart. So the intensity of infrared reflected by the skin (and thus transmitted to the phototransistor) rises and falls with each heartbeat.

CONTACT DEBOUNCER

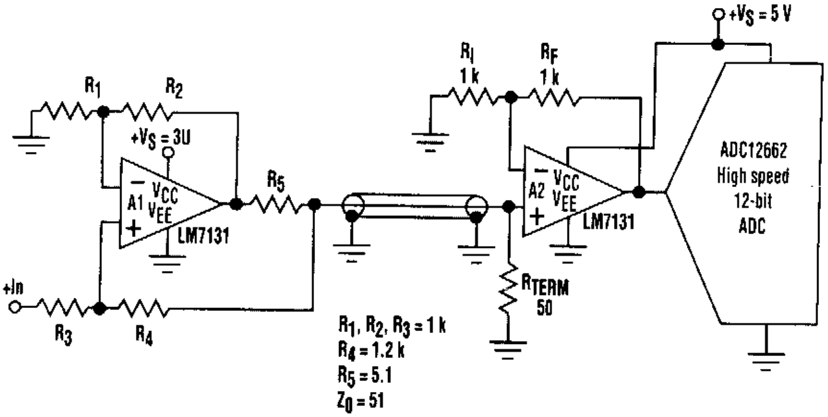


ELECTRONICS NOW

Fig. 52-28

A contact debouncer using a Schmitt trigger, such as a TTL7414, provides a "clean" pulse from a switch contact closing.

POSITIVE FEEDBACK CABLE TERMINATOR

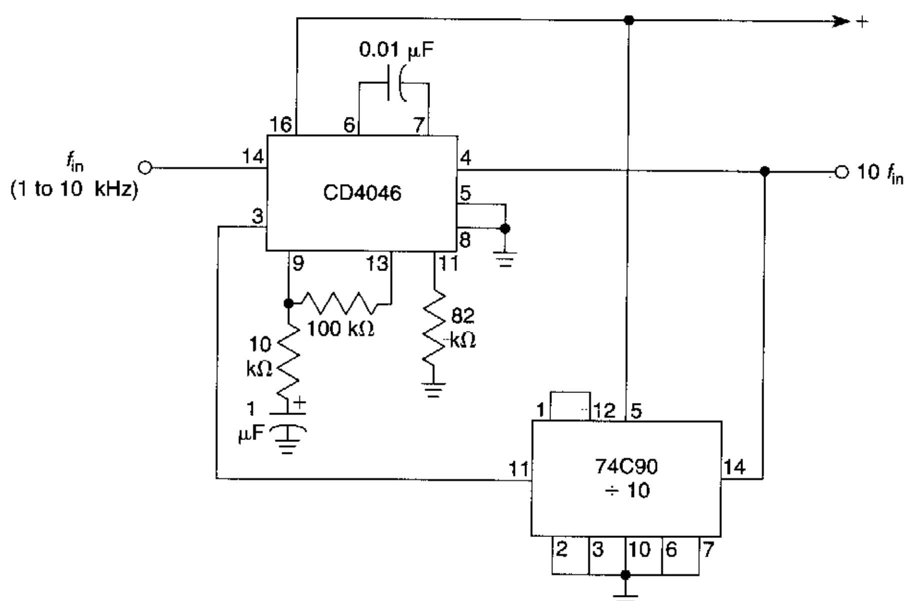


ELECTRONIC DESIGN

Fig. 52-29

Positive feedback along with a series output resistor can provide a controlled output impedance from an op-amp circuit. The circuit is useful when driving coaxial cables that must be terminated at each end in their characteristic impedance, which is often 50 Ω. Adding a 50-Ω series resistor on the op amp's output obviously reduces the available signal swing.

×10 FREQUENCY MULTIPLIER

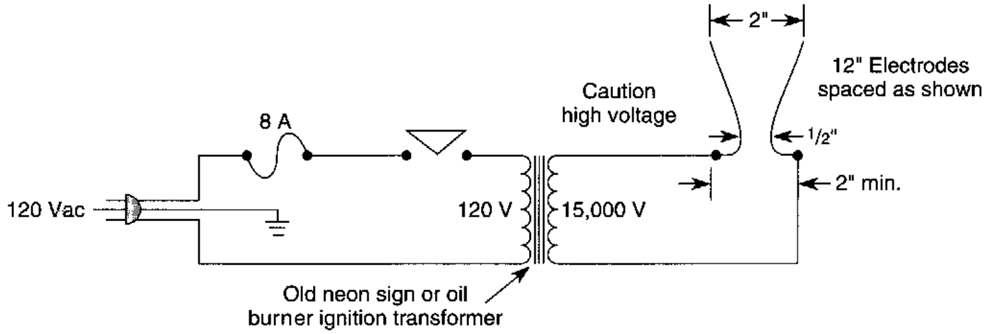


WILLIAM SHEETS

Fig. 52-30

In this circuit, the CD4046 is set up so that the V_{CO} operates at 10 to 100 kHz. The output pin (4) is fed back to a ÷10 counter. When the input frequency is 1/10 the output, lockup will occur.

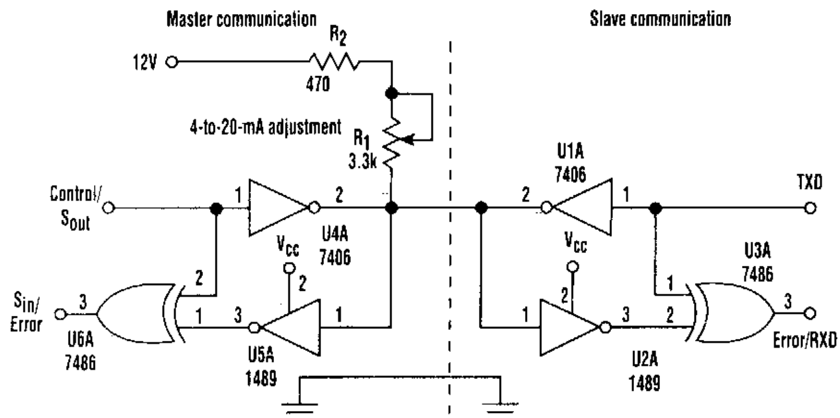
JACOB'S LADDER



WILLIAM SHEETS **Fig. 52-31**

A "Jacobs Ladder" can be made from an old neon sign or oil burner ignition transformer. A rating of 12 to 15 kV at 20 to 30 mA will be adequate. Make sure to mount the electrodes to a pair of insulators, at least 2" apart, and bent and spaced, as shown. The ladder should be enclosed in a clear plastic housing to prevent accidental contact with the high voltage and to ensure a stable arc. Vent holes should be placed top and bottom to allow gases to escape.

MASTER-SLAVE DEVICE ERROR CHECKER

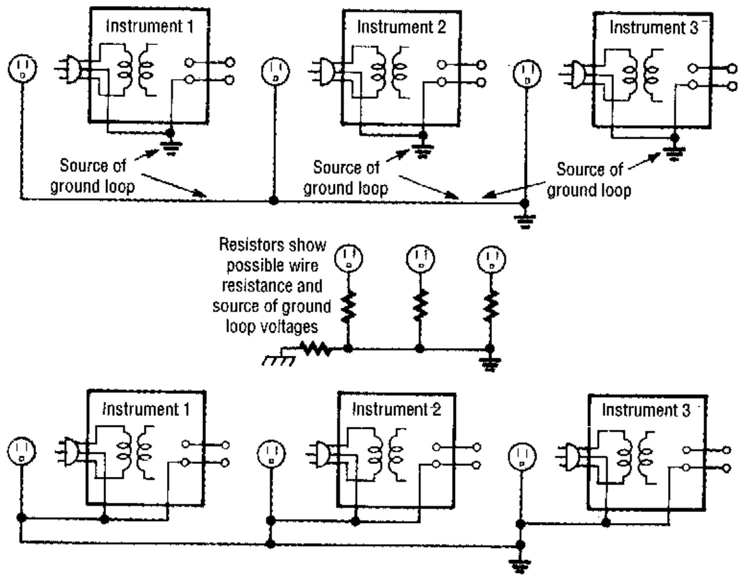


ELECTRONIC DESIGN **Fig. 52-32**

An error-check mechanism introduced into master and slave communication devices can indicate mismatches when both the master and slave start sending data simultaneously. The error flag goes high and indicates a mismatch in the data.

The master is the one that can interrupt the communication from a device at the other end and force it to listen. It does this forcing a low voltage level over the communication line by raising the control line to a high level. This inhibits data flow over the lines from the slave device. As a result, the slave turns into a listen mode (not a hardware feature, but rather incorporated in the software). The slave device can transmit the data after communication from the master device ceases.

GROUND LOOP PREVENTER

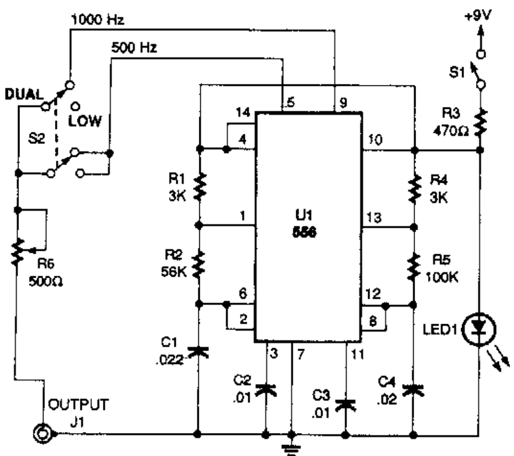


ELECTRONICS NOW

Fig. 52-33

Ground loops are caused by improper grounding. Ground-loop voltages can interfere with test measurements because the voltages in a ground loop can be larger than the signals you're trying to measure. To prevent ground loops, use two wire plugs to provide the line power to the test instruments and a separate wire to bring the input grounds of the instruments to a common ground.

DUAL TONE GENERATOR FOR AUDIO SERVICING

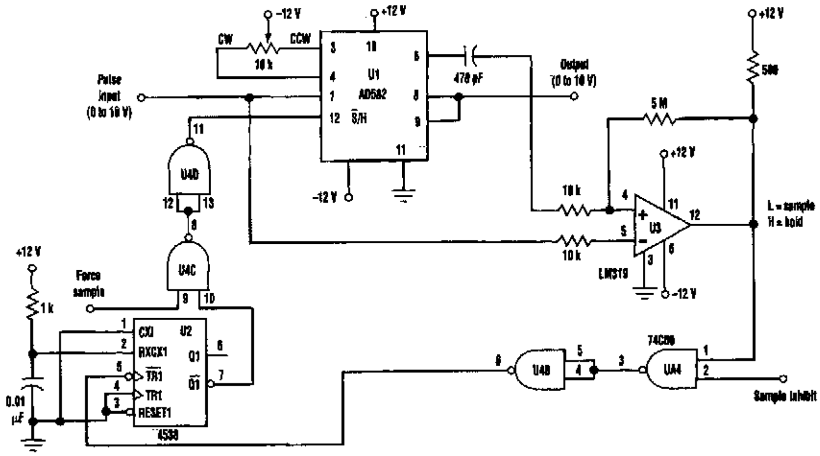


POPULAR ELECTRONICS

Fig. 52-34

This dual-tone generator can insert a distinctive tone in the audio section of a circuit under test. That way, you can work your way back from the speaker, stage-by-stage, to locate a faulty section.

DIODELESS PEAK-HOLD CIRCUIT



ELECTRONIC DESIGN *Fig. 52-35*

The input pulse is fed into the sample-and-hold amplifier (an inexpensive AD582), as well as the comparator U3. The SHA's output also is fed into the comparator. If the input pulse is higher in amplitude than the SHA's output, the comparator output goes low and the 4538 one-shot produces a 10-µs pulse that is fed back to cause the SHA to sample and then hold the voltage. Subsequent input voltages that are less than the held value won't cause the one-shot to fire again.

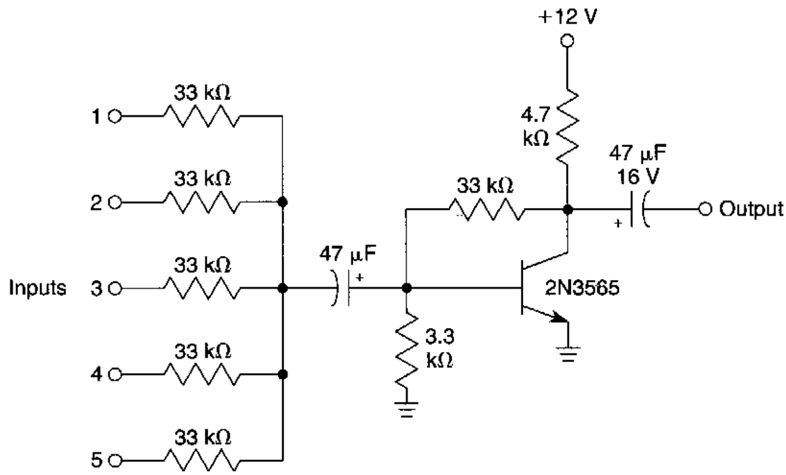
Gates U4A and U4B are used to inhibit the sampling when necessary. Gates U4C and U4D, at the one-shot's output, can force the AD 582 into the sample mode. This feature is useful to reset the output to zero by forcing a sample when the input to the AD582 is zero. The polarity of the peak-hold circuit can be easily changed from positive-to-negative peak hold by reversing the inputs of the comparator.

53

Mixer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

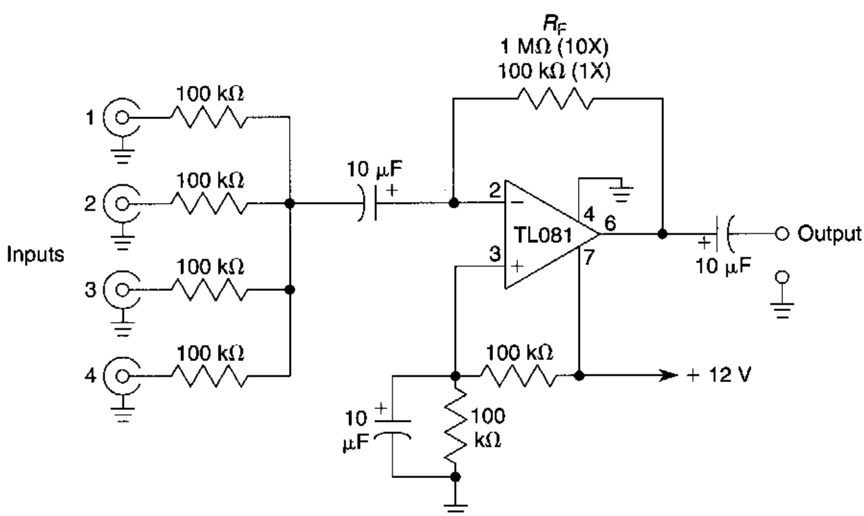
SIMPLE AUDIO MIXER



WILLIAM SHEETS *Fig. 53-1*

A single transistor is used as an audio mixer, the transistor serving as a feedback amplifier.

OP-AMP AUDIO MIXER



WILLIAM SHEETS *Fig. 53-2*

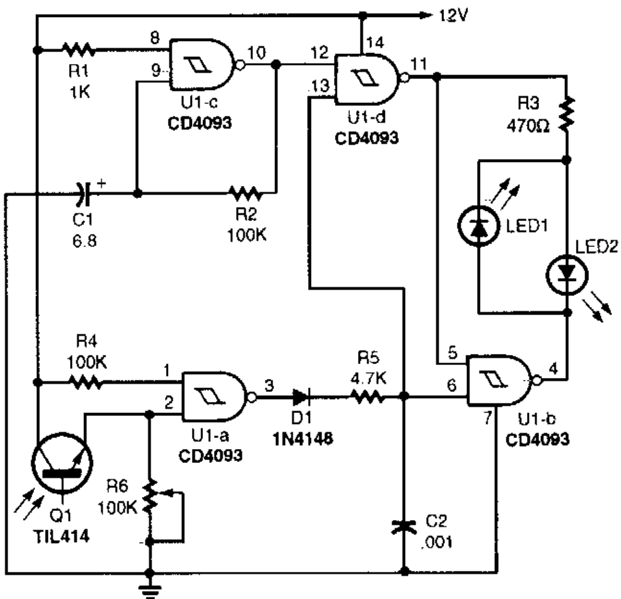
This circuit will mix several audio signals to a common output. R_F can be made 1 MΩ for 10 × (20 dB) or 100 kΩ for unity gain.

54

Model and Hobby Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

MODEL RAILROAD CROSSING FLASHER

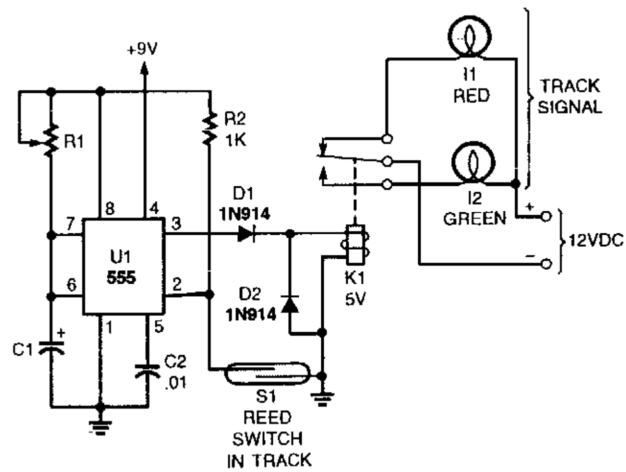


POPULAR ELECTRONICS *Fig. 54-1*

Gate U1-c is set up as an oscillator whose frequency is determined by C1 and R1. Gates U1-b and U1-d are set up as an RS flip-flop that is gated on by U1-a. Gate U1-a in conjunction with Q1 operates as the control gate for the flip-flop. Components D1, C2, and R5 act as a delay circuit to compensate for any light getting through the gaps between cars as they pass over the phototransistors. The light-emitting diodes are connected so that they operate alternately, depending on the outputs of U1-d and U1-b.

Basically, R6 is adjusted so that ambient room-light striking Q1 (and any other phototransistors connected in series) keeps the output of U1-a at pin 3 low. When a car passes over the phototransistor, which is installed between ties in the track, pin 3 goes high, allowing a high to be placed on pins 5 and 13. That allows the high output of U1-c at pin 10 to enable pin 12, which in turn allows pin 11 to go low. That makes a complete path for LED2 to operate. When pin 10 goes low, pin 11 goes high. That makes pin 5 high, and thus, enables pin 4 to go low and completes the circuit for LED1. That alternates the LEDs, which are installed in a railroad-crossing signal.

MODEL RAILROAD TRACK CONTROL SIGNAL



RED-LAMP ON TIME (SECONDS)		
R1 (KILOHMS)	C1 = 10μF	C1 = 100μF
100	2	16
220	3	32
470	6	70
1000	15	175

POPULAR ELECTRONICS *Fig. 54-2*

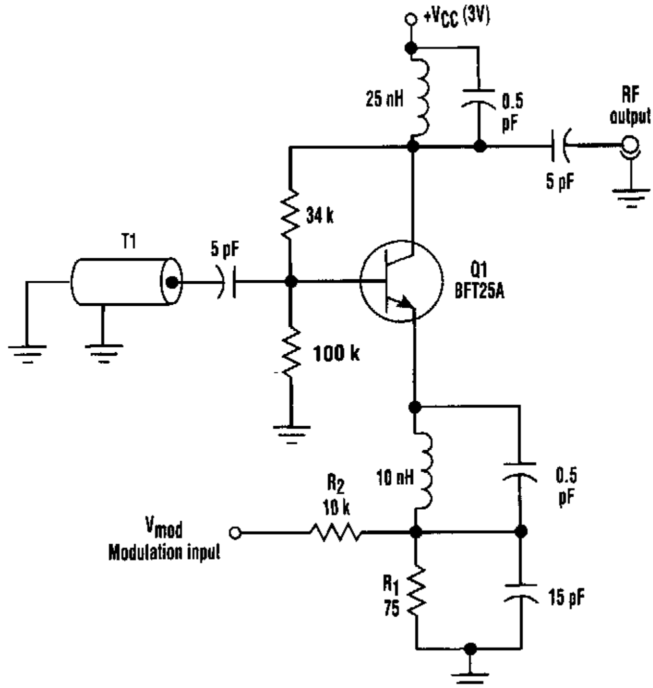
When a train passes S1 (a red switch), a small magnet glued to the underside operates S1 and causes U1 to generate a pulse, activating relay K1 and changing the signal from green to red. After a time determined by R1 and C1 (see table), the relay de-energizes and the signal goes back to green.

55

Modulator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

VARACTORLESS HF MODULATOR



ELECTRONIC DESIGN

Fig. 55-1

Traditionally, high-frequency oscillators are frequency-modulated by using a varactor. However, varactors usually require a large voltage change to achieve a reasonable capacitance change—a problem in many battery-powered systems.

Such a problem can be overcome by employing base-charging capacitance modulation. Resistor R1 establishes Q1's current, and R2 allows control of the collector bias current by V_{mod}. The transmission line (T1) in the negative resistance-type oscillator determines the frequency of oscillation. T1 is a high-quality, low-loss, ceramic coaxial shorted quarter-wave transmission line. Under proper terminal impedances, a negative resistance is "seen" at Q1's base. T1 reacts with this negative resistance to produce sustained oscillations.

Frequency modulation is accomplished by changing Q1's collector bias current and thus changing Q1's base-charging capacitance. This effect is "seen" at Q1's base and causes a frequency shift in the resonators quarter-wave node.

MODULATOR FOR VIDEO

Sources: SAWFs

Crystal Technology, Inc.
1035 E. Meadow Circle
Palo Alto, CA 94303

Kyocera International, Inc.
8611 Balboa Ave.
San Diego, CA 92123

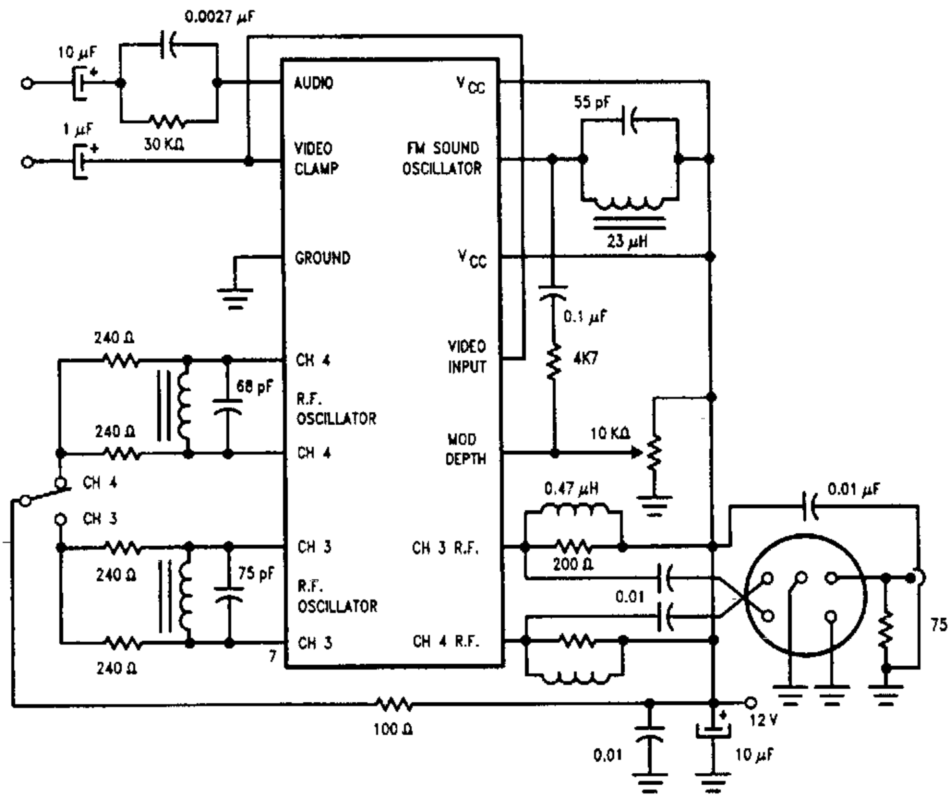
MuRata Corp. of America
1148 Franklin Rd. S.E.
Marietta, GA 30067

CRYSTALS

Saronix
4010 Transport at San Antonio Rd.
Palo Alto, CA 94303

COILS

Toko America, Inc.
5520 W. Touhy Ave.
Skokie, Ill. 60077



NATIONAL SEMICONDUCTOR *Fig. 55-2*

This circuit uses an LM2889 and a saw filter for use as a TV modulator.

DIGITAL PULSE-WIDTH MODULATION CIRCUIT

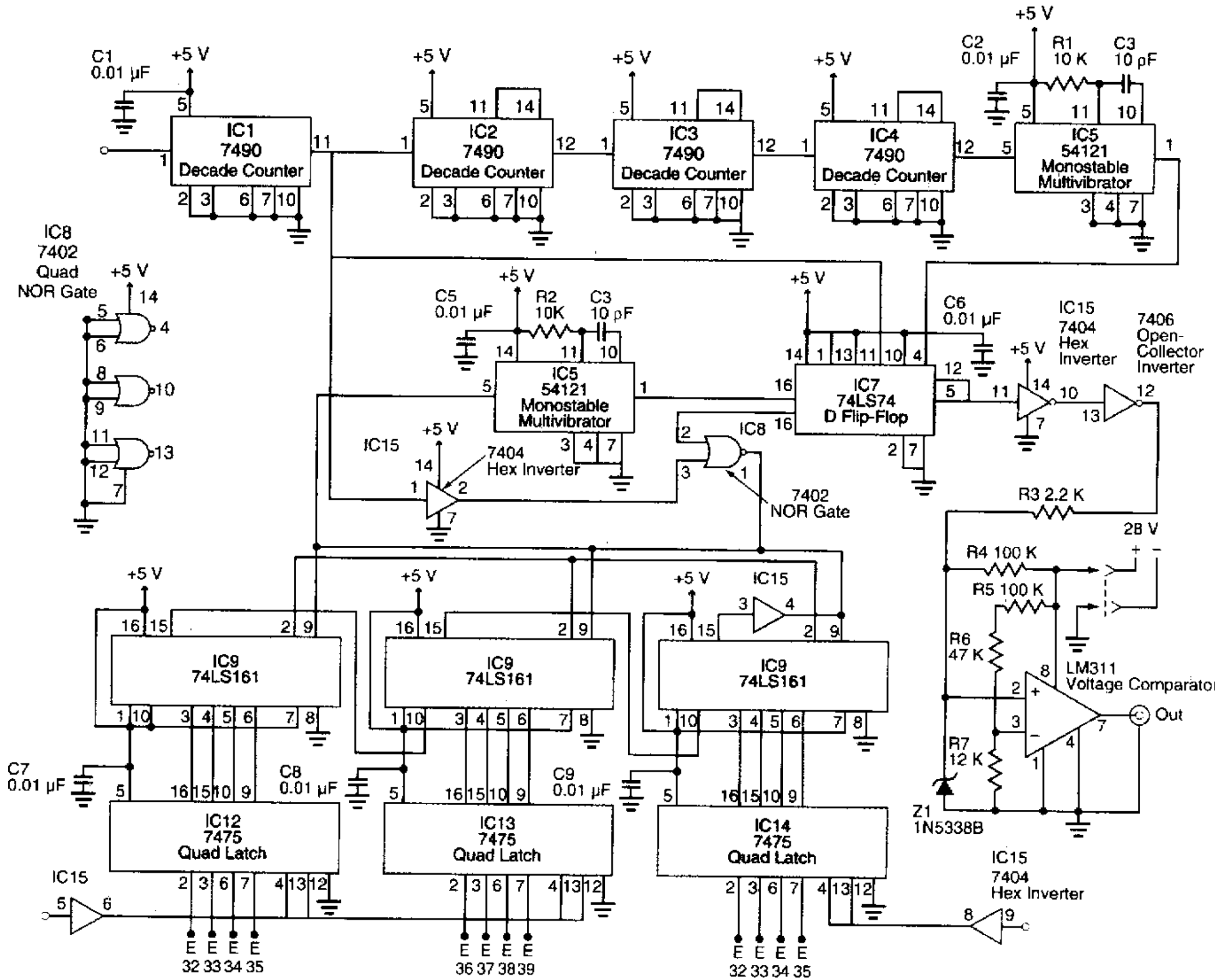


Fig. 55-3

DIGITAL PULSE-WIDTH MODULATION CIRCUIT (*Cont.*)

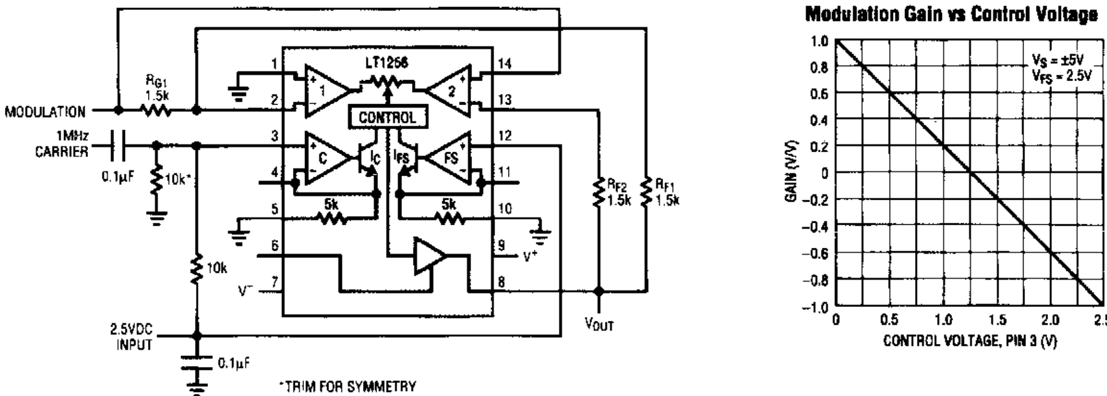
This circuit incorporates low-power Schottky transistor/transistor-logic (TTL) devices in critical high-speed parts. The 5-MHz clock signal is converted to a 1-MHz signal by a 7490 decade counter configured to divide by 5. The 1-MHz signal is sent, via a clock synchronizer, to a three-stage binary divider that consists of three cascaded 74LS161 binary dividers. The synchronizer consists of a 74LS74 D flip-flop, a 7404 inverter gate, and a 7402 NOR gate. The dividers are programmed from the STD bus by means of three 7475 quad latches; this makes it possible to program the frequency division from 1 to 4096 (12 bits).

The 1-MHz signal is also divided by 1000 by use of three cascaded 7490 decade counters, each configured to divide by 10; this provides a 1-kHz signal, which is sent to a 54121 monostable multivibrator configured to provide a 0.1-μs pulse, bombarded with an ion beam source in preparation for the materials about to be deposited. While the surface is bombarded with an ion beam, an electron beam source is activated so that a layer of fused silica is vapor-deposited to a total desired thickness value (typically, 1 micron or 10,000 Å). The layer of fused silica serves as a surface stabilization layer for the next step.

A metal mask with an aperture in the specified pattern of the sensor film is placed on the surface at the specified sensor location. The surface area exposed through the mask is cleaned by ion-beam bombardment for a predetermined time. Then as the bombardment continues, a metal (typically, nickel, platinum, and/or palladium) is vapor-deposited through the mask from the electron-beam source to form the sensor film. Deposition is continued until the thickness of the film reaches the value specified in the particular sensor design. A representative value for a nickel sensor film is 2500 Å.

Next, a pattern for thin film leads is defined by taping directly on the surface of the model with Kapton (or equivalent) polyimide tape. The thin film leads are fabricated by a combination of ion-beam bombardment and electron-beam vapor deposition like that used to deposit the sensor film. The metal vapor deposited in this step is typically copper, gold, or aluminum. A typical thickness for copper leads on the nickel sensor film is about 10,000 Å.

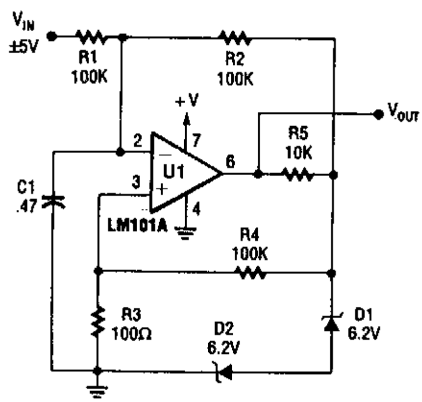
FOUR-QUADRANT MULTIPLIER AS DSB MODULATOR



LINEAR TECHNOLOGY

Fig. 55-4

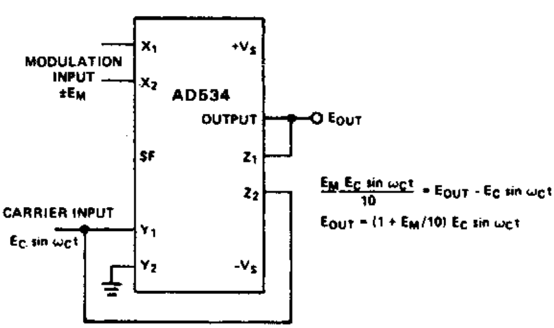
PULSE-WIDTH MODULATOR



POPULAR ELECTRONICS

Fig. 55-5

LINEAR (AM) AMPLITUDE MODULATOR

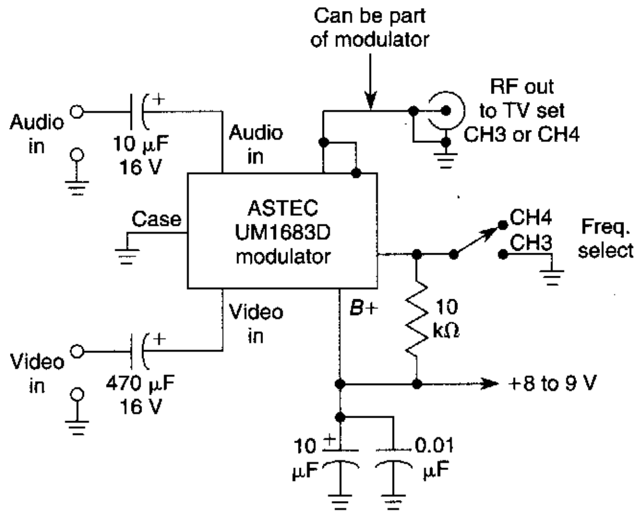


This is a very simple amplitude modulator. It makes use of the Z2 terminal to add the carrier directly to the output, thus bypassing the multiplier for zero modulation input. It has the advantage of allowing operation from a differential modulation input.

ANALOG DEVICES

Fig. 55-6

VIDEO MODULATOR HOOKUP



WILLIAM SHEETS *Fig. 55-7*

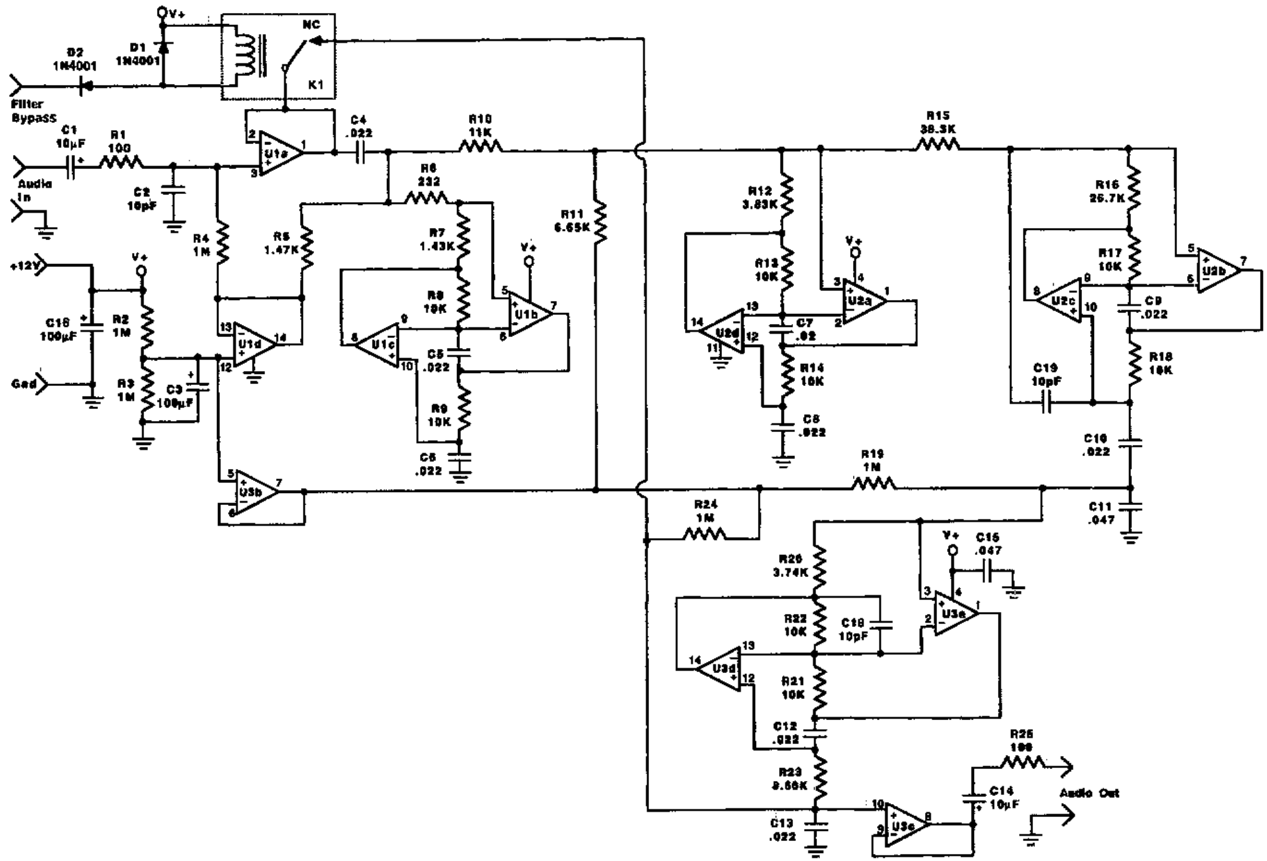
This circuit uses an ASTEC UM1683D, but it is typical of many RF video modulators used in VCRs and satellite receivers.

56

Morse Code Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

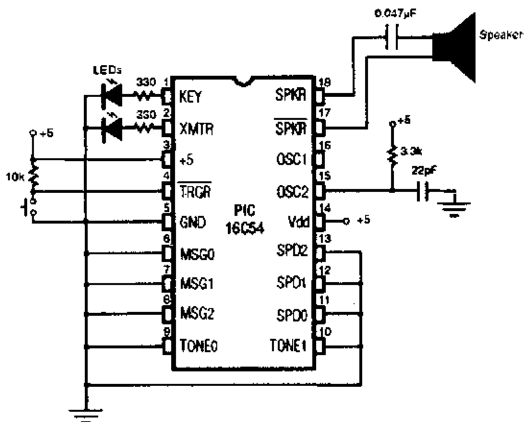
ACTIVE CW AUDIO FILTER



73 AMATEUR RADIO TODAY *Fig. 56-1*

The audio filter shown has a bandpass of 200 Hz centered on 700 Hz. Resistors are 1% tolerance and capacitors should be 5% tolerance.

MORSE MESSENGER



MSG2	MSG1	MSG0	Message No.
0	0	0	0
0	0	1	1
0	1	0	2
0	1	1	3
1	0	0	4
1	0	1	5
1	1	0	6
1	1	1	7

TONE1	TONE0	Sidetone (Hz)
0	0	400
0	1	525
1	0	650
1	1	775

SPD2	SPD1	SPD0	Speed (WPM)
0	0	0	7
0	0	1	10
0	1	0	15
0	1	1	20
1	0	0	25
1	0	1	30
1	1	0	35
1	1	1	40

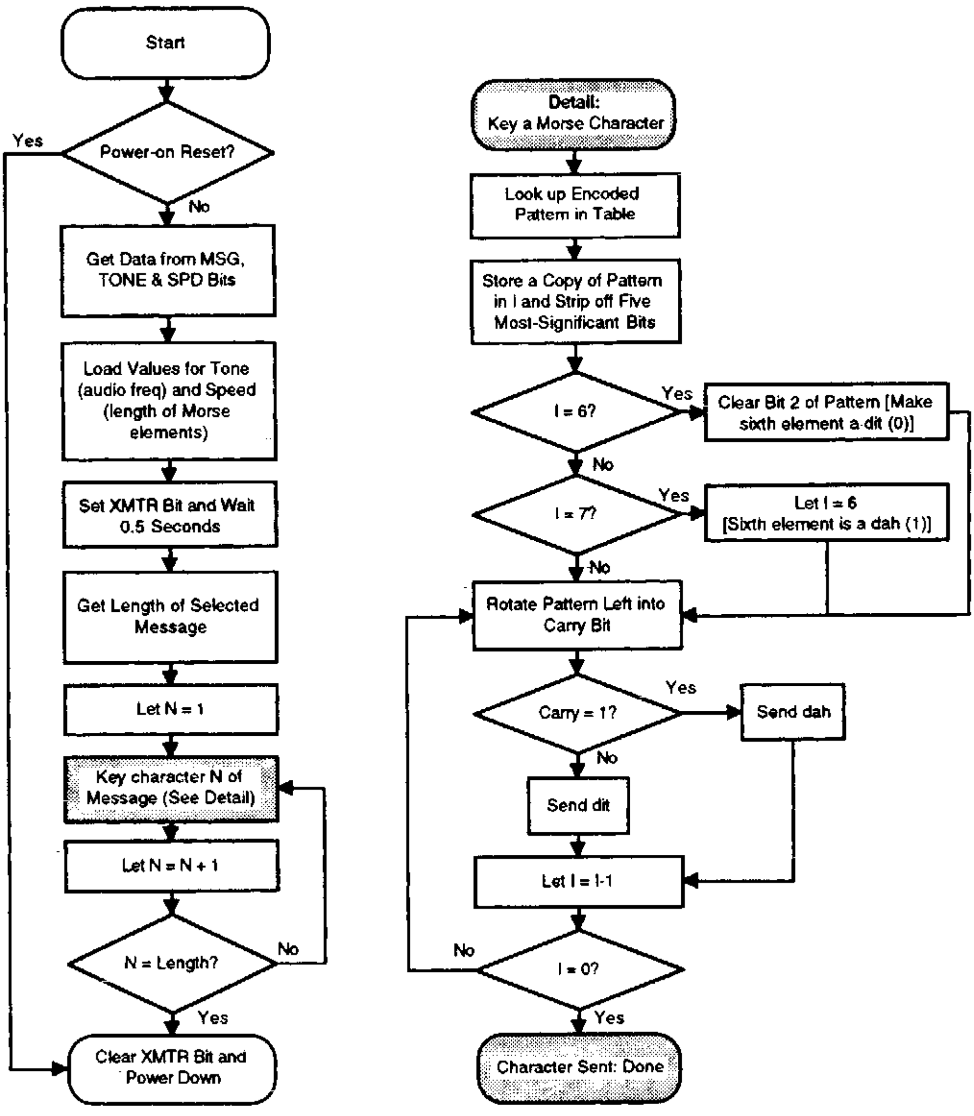
Set up as shown, the chip will play message no. 0 at 7 words per minute, lowest sidetone frequency (400 Hz). LEDs indicate oscillator warmup (XMTR) and keying (KEY).

Simple hook-up diagram for the Morse Messenger chip. The table indicates the range of messages, sidetones, and keying speeds.

73 AMATEUR RADIO TODAY

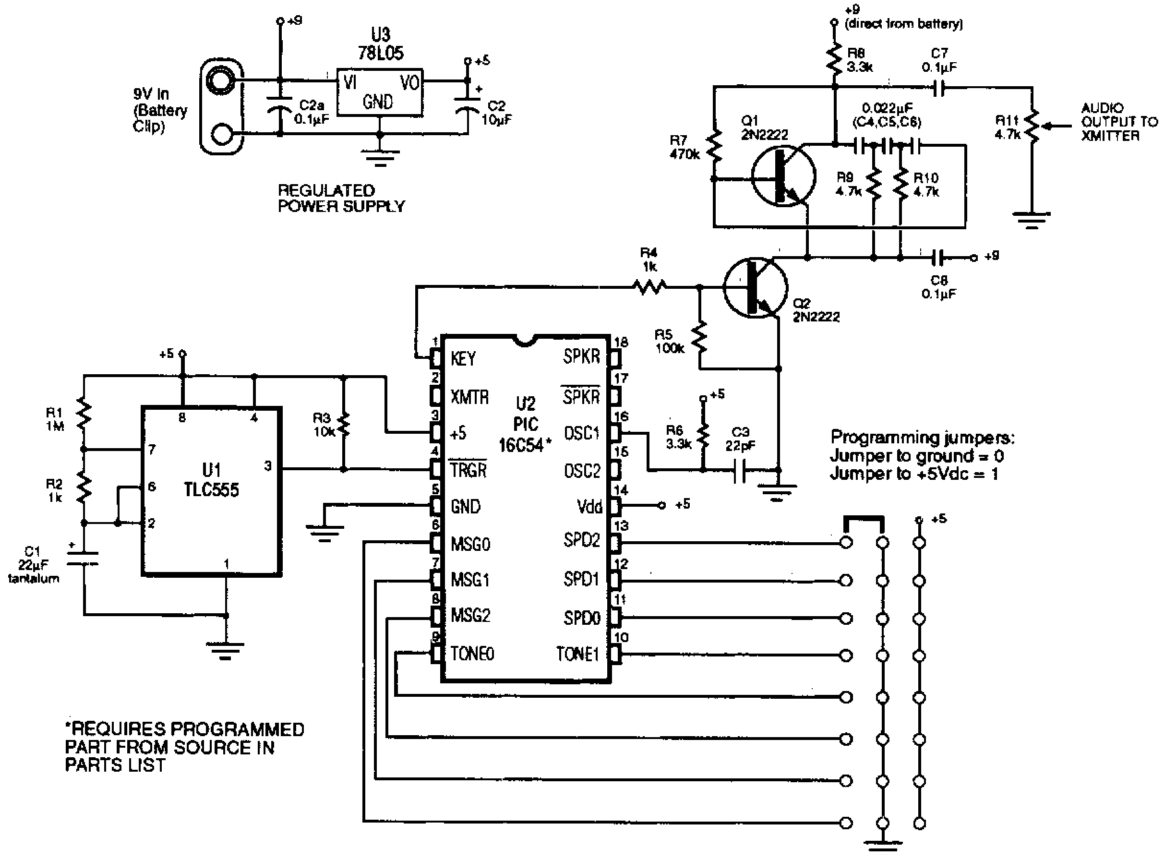
Fig. 56-2

This keyer uses a PIC16C54 micro-controller to generate a Morse code message. The microcontroller must be programmed to suit users call IC or desired message.



Logic of the Morse Messenger's program. This algorithm can be adapted to other devices with the help of the Morse encoding table.

CW IDENTIFIER WITH SINE-WAVE AUDIO OUTPUT

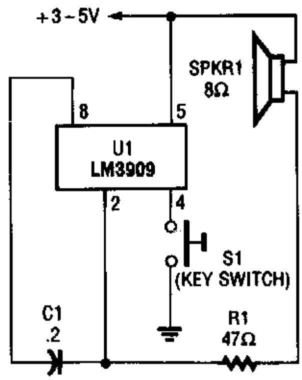


73 AMATEUR RADIO TODAY

Fig. 56-3

This identifier can be used to drive a hidden transmitter in a radio "fox hunt" activity, where the object is to locate a hidden transmitter.

SIMPLE CODE PRACTICE OSCILLATOR



POPULAR ELECTRONICS ***Fig. 56-4***

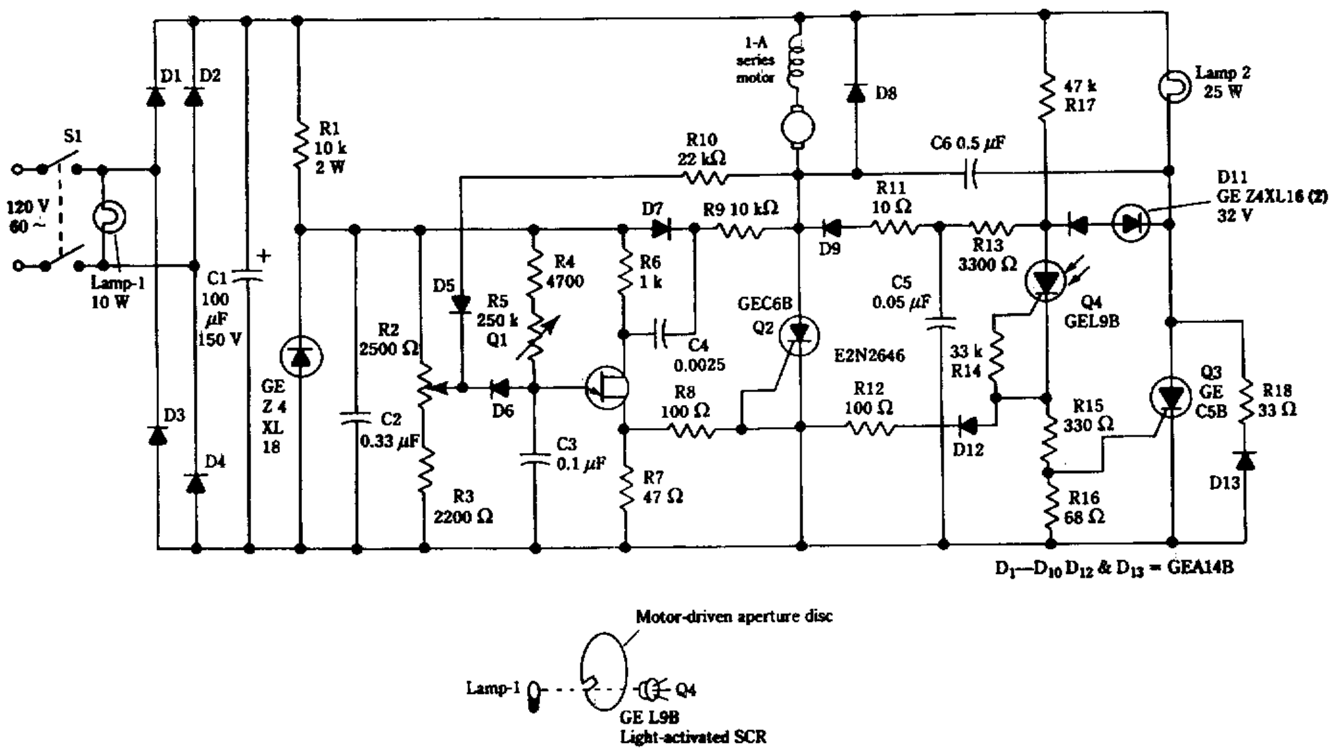
With only a minor circuit change, the basic LM3909 oscillator configuration can be turned into a code-practice oscillator.

57

Motor-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

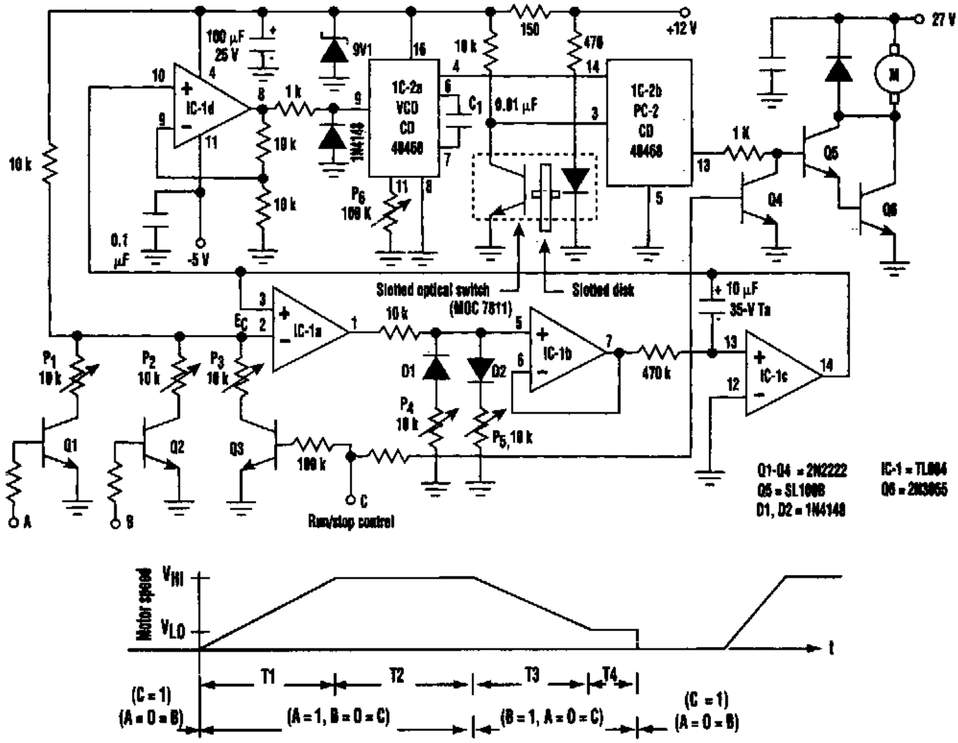
PRECISE dc MOTOR SPEED CONTROLLER



McGRAW-HILL

Fig. 57-1

A series dc motor can be made to have the same characteristics as an ac synchronous motor using this circuit. This control technique is useful where a constant motor speed is needed.



This motor velocity profile is provided by the controller. Note the smooth acceleration and deceleration profiles and the constant motor speed between the two ramps.

ELECTRONIC DESIGN **_Fig. 57-2_**

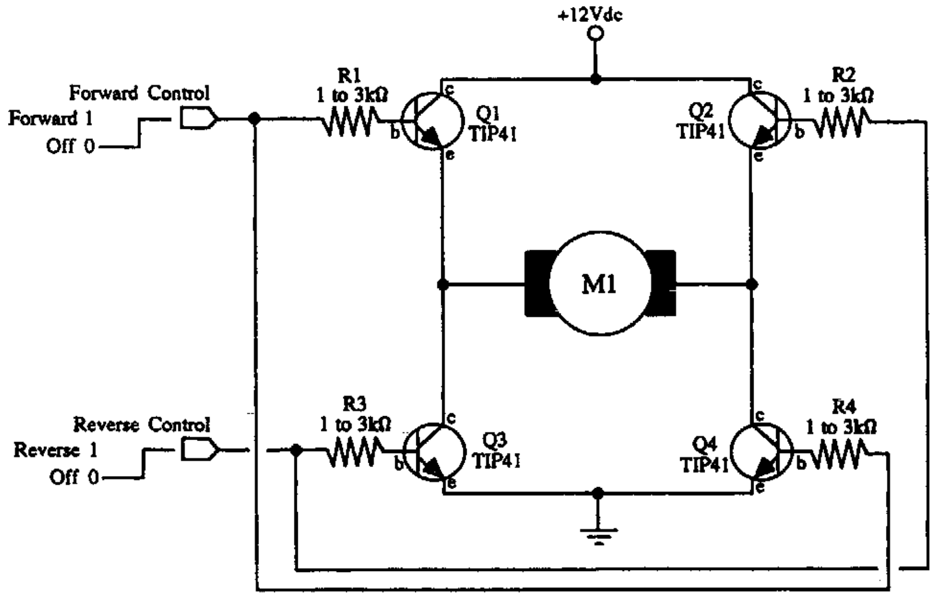
This circuit can control dc motors used in machines that pull materials, such as wire, yarn, film, etc., from the supply rolls and rewind them onto smaller rolls. Its motor velocity profile is shown in the figure.

A lightweight disk with at least 32 slots is mounted on the motor shaft. It works in a slotted optical switch (MOC7811) to sense the motor speed. Phase comparator 2 (PC-2) of the phase-locked loop (IC-CD 4046B) compares the frequency (f_v) at the output of the VCO with the pulse rate (f_m) at the optical switch output. The PC-2 output drives the motor via the transistors when $f_v > f_m$, and removes the supply to the motor when $f_m > f_v$. The drive system quickly reaches an equilibrium condition when $f_m = f_v$.

Op amps IC-1a, 1b, and 1c form a tracking integrator whose output always smoothly reaches and remains at a voltage equal to the command voltage (E_c) presented at the inverting input of op-amp IC-1a. When the digital control inputs are set to ($A = 1, B = 0 = C$), the integrator generates a positive slope ramp that sweeps the VCO frequency and, thus, accelerates the motor. The acceleration rate and the constant speed V_{HI} can be adjusted by presets P4 and P1, respectively.

Similarly, when the inputs are set to ($A = 0 = C, B = 1$), the integrator generates a negative slope ramp that decelerates the motor. Presets P5 and P2 can be adjusted to set the required deceleration and the constant speed V_{LO}, respectively.

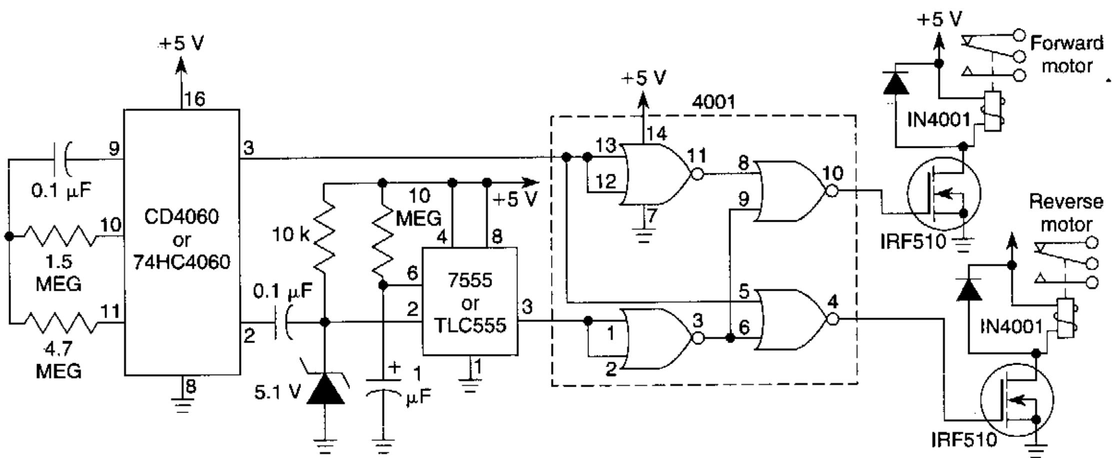
MOTOR DIRECTION CONTROL USING DISCRETE TRANSISTORS



McGRAW-HILL *Fig. 57-3*

For best operation, mount the transistors on heatsinks. The transistors specified are fine for small hobby motors, or up to about 6 volts dc and between 800 and 1000 mA. M1 is a small hobby motor.

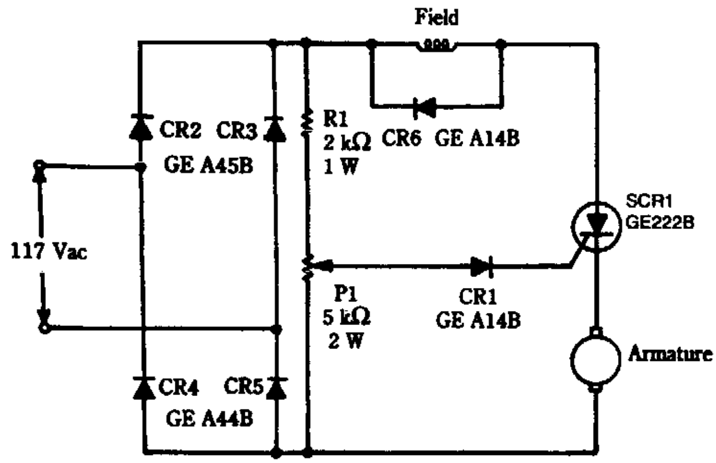
LONG TIME-DELAY MOTOR-CONTROL CIRCUIT



ELECTRONICS NOW *Fig. 57-4*

Circuit controls forward and reverse motors. Every hour, one motor runs for 10 seconds.

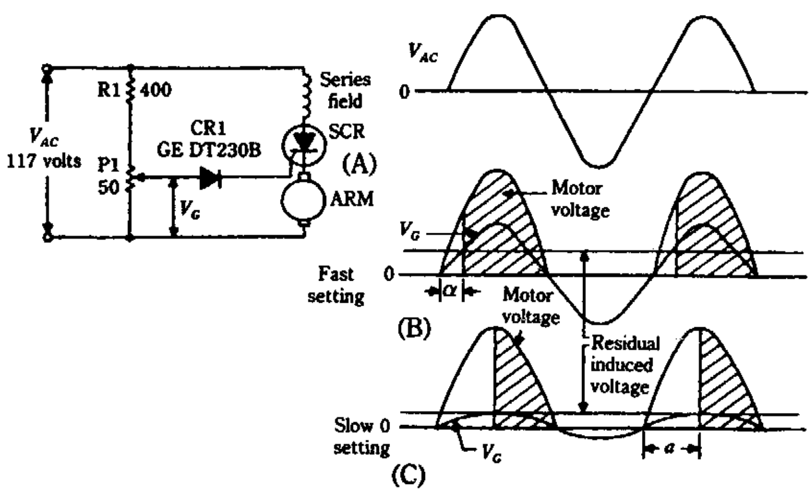
FULL-WAVE SPEED CONTROL FOR MOTORS



McGRAW-HILL *Fig. 57-5*

A bridge rectifier provides pulsating dc to a universal motor, and the SCR is used as a phase-controlled switch. This circuit allows smoother operation of the motor at low speeds.

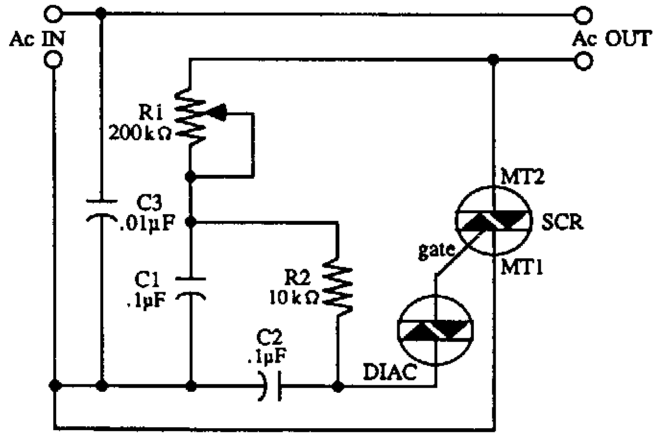
SCR MOTOR SPEED CONTROL



McGRAW-HILL *Fig. 57-6*

An SCR is used in a phase-control type application to supply a variable pulsating dc voltage to a motor.

TRIAC MOTOR-CONTROL CIRCUIT

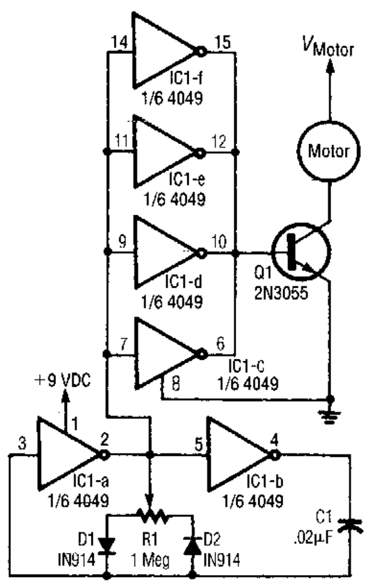


McGRAW-HILL

Fig. 57-7

An SCR-controlled ac motor control circuit. This is a full-wave circuit and is best used when the load remains constant.

LOW-VOLTAGE dc MOTOR-SPEED CONTROLLER

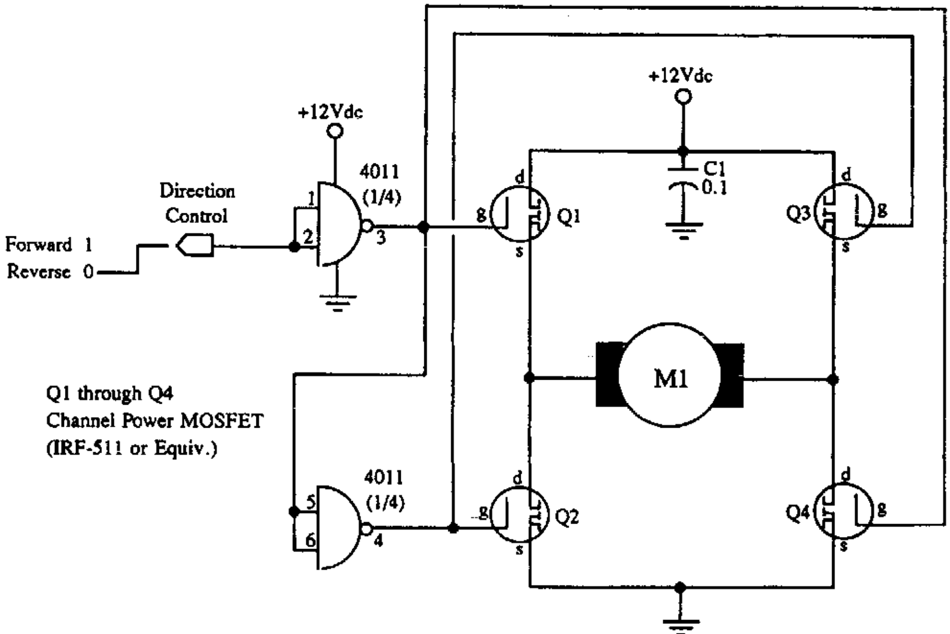


ELECTRONICS NOW

Fig. 57-8

This circuit varies the duty cycle, rather than the voltage. The two diodes control the positive and negative halves of the capacitor's charging cycle.

MOTOR DIRECTION CONTROL



McGRAW-HILL *Fig. 57-9*

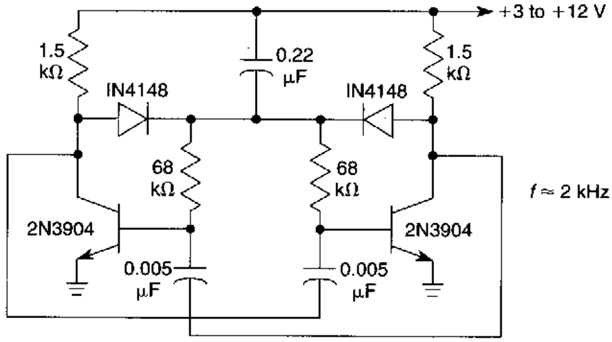
M1 is a small hobby dc motor.

58

Multivibrator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

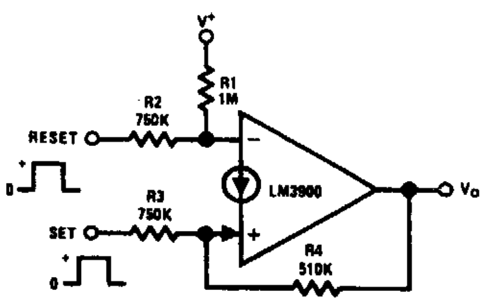
ASTABLE MULTIVIBRATOR WITH STARTING NETWORK



WILLIAM SHEETS *Fig. 58-1*

This circuit will start with a slowly rising supply voltage waveform.

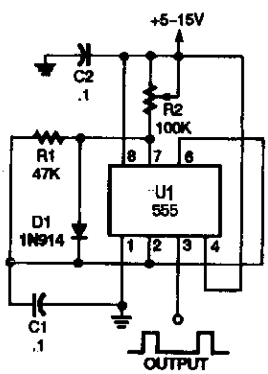
BISTABLE MULTIVIBRATOR



Positive feedback is provided by resistor R4, which causes the latching. A positive pulse at the "set" input causes the output to go high and a "reset" positive pulse will return the output to essentially 0 Vdc.

NATIONAL SEMICONDUCTOR *Fig. 58-2*

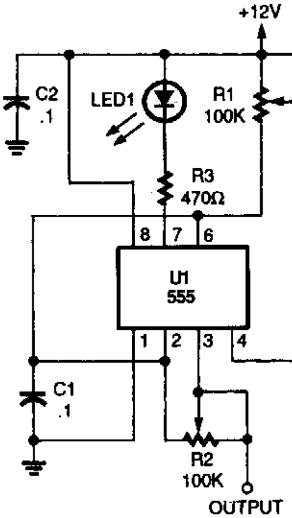
ASTABLE



At the instant that power is applied to the 555 astable, timing capacitor C1 is initially discharged, causing the output of the chip output at pin 3 to be high. Once C1 has charged to about ⅔ of the supply voltage, its output goes low, and the discharge transistor turns on, draining the charge on C1.

POPULAR ELECTRONICS *Fig. 58-3*

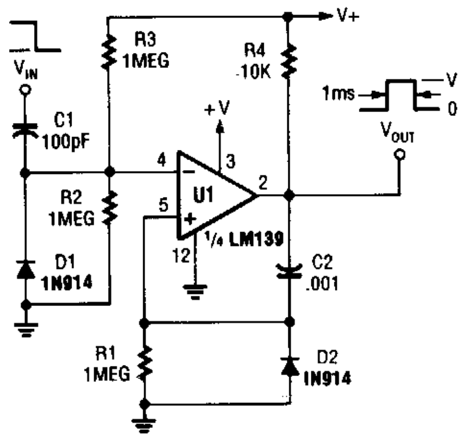
ASTABLE WITH VARIABLE PULSE WIDTH



POPULAR ELECTRONICS *Fig. 58-4*

This produces a positive variable width pulse and has a symmetry control. R1 and R2 control the pulse width and symmetry.

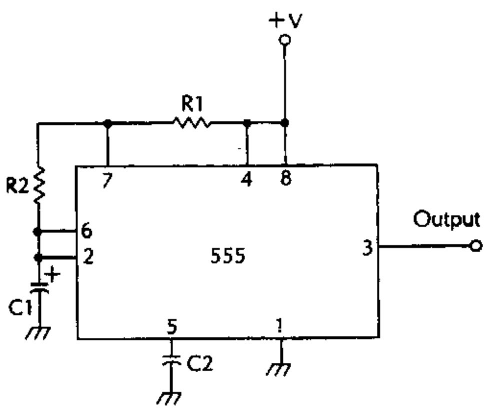
ONE-SHOT MULTIVIBRATOR



POPULAR ELECTRONICS *Fig. 58-5*

An LM139 section can be used as a one shot.

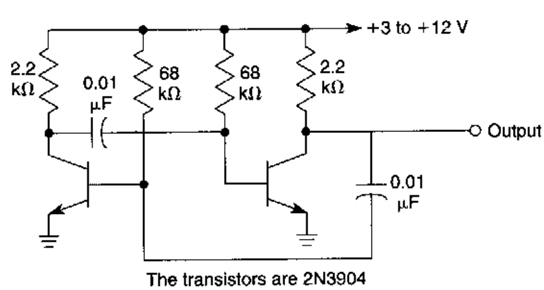
BASIC 555 ASTABLE MULTIVIBRATOR



McGRAW-HILL *Fig. 58-6*

$$F = 1.44$$
$$(R_1 = 2R_2)\ C_1$$

ASTABLE MULTIVIBRATOR



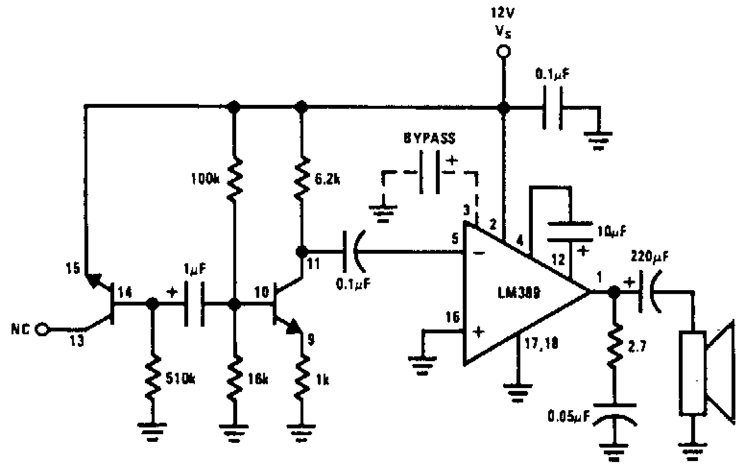
WILLIAM SHEETS *Fig. 58-7*

59

Noise Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

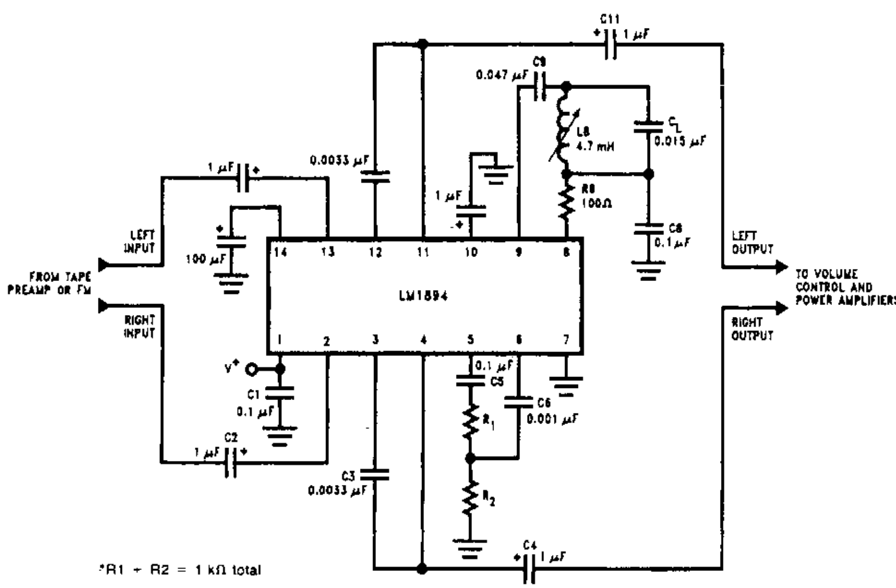
NOISE GENERATOR



NATIONAL SEMICONDUCTOR *Fig. 59-1*

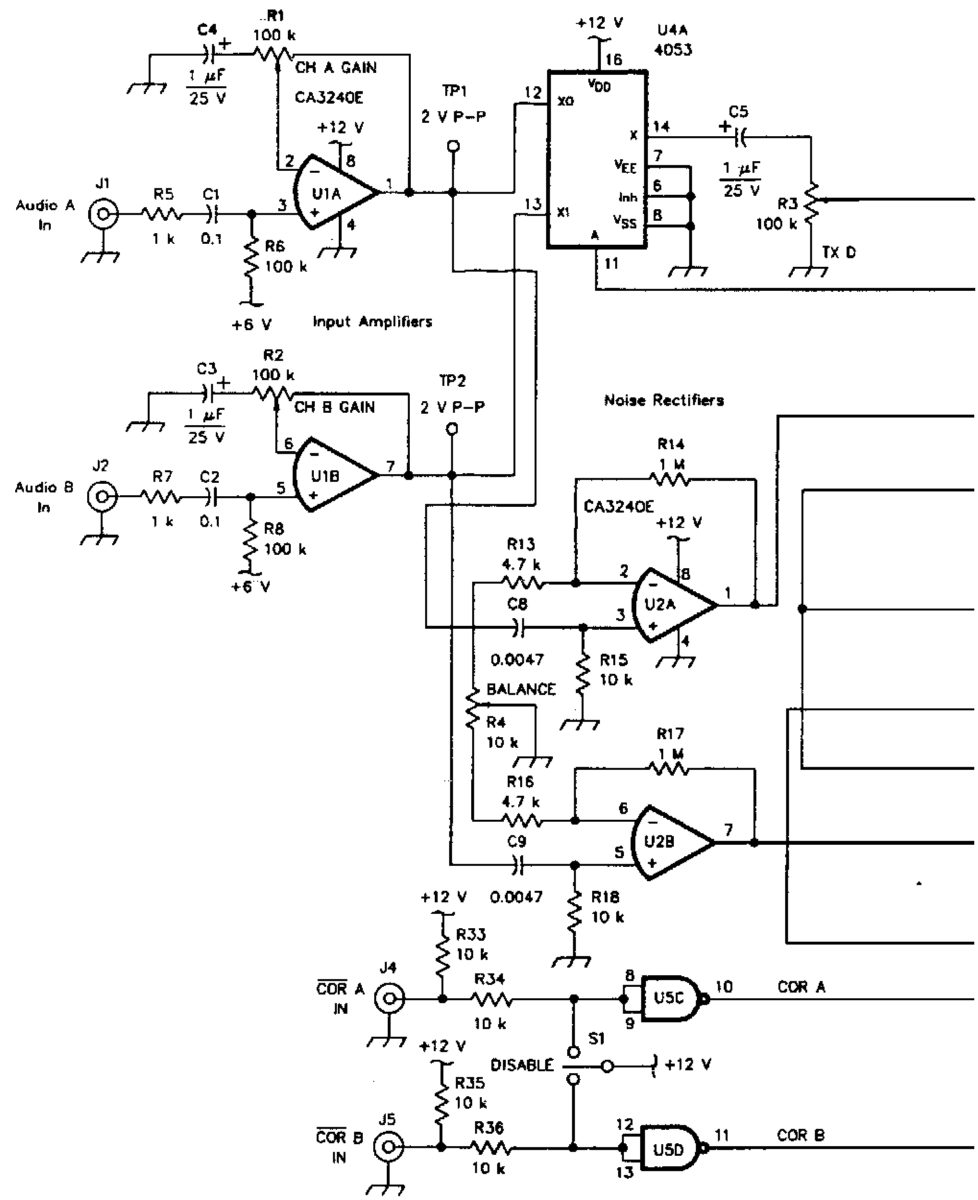
This noise generator uses a Zener diode. The transistors are part of the LM389.

DOLBY NOISE-REDUCTION CIRCUIT



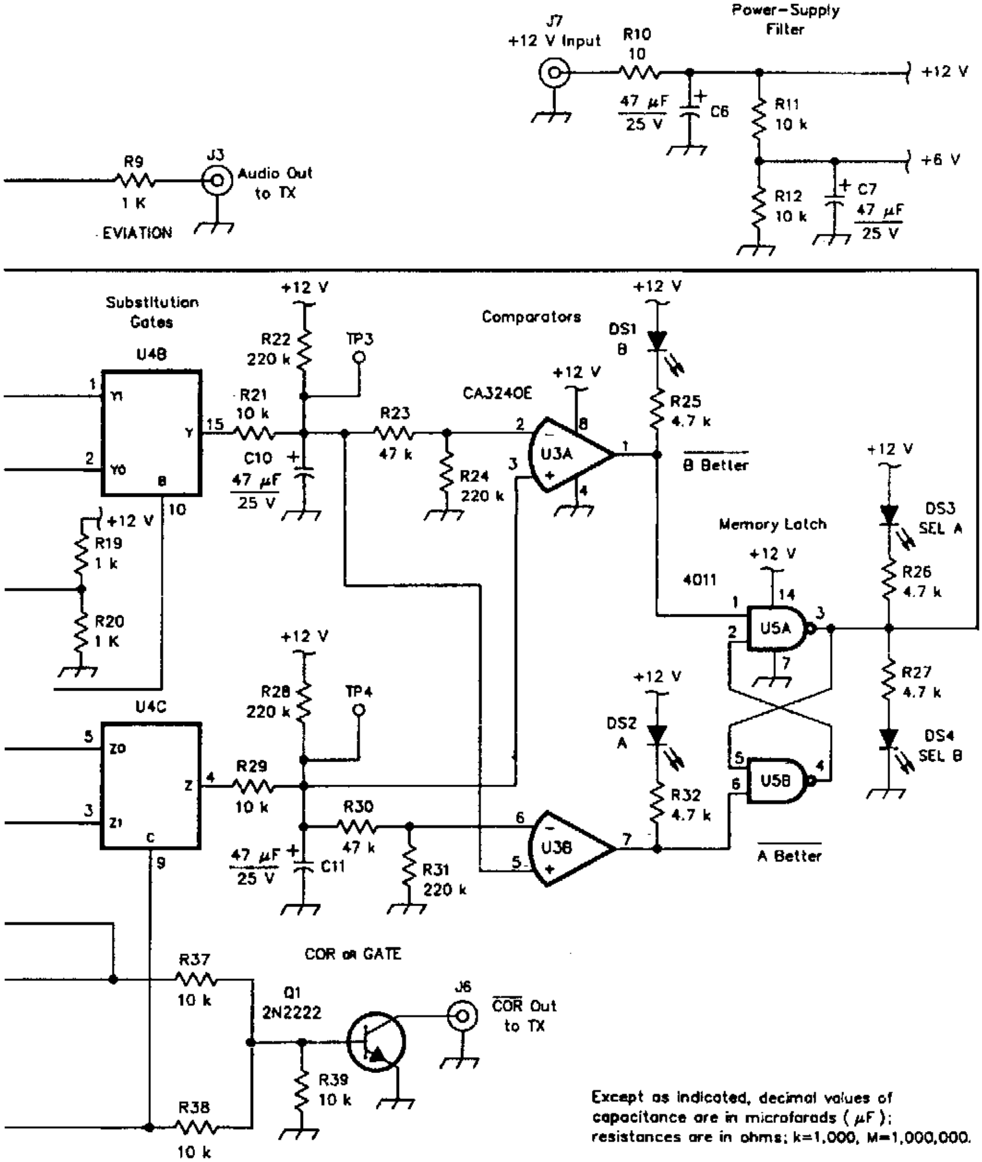
NATIONAL SEMICONDUCTOR *Fig. 59-2*

AUDIO NOISE-BASED VOTING CIRCUIT



QST

The purpose of this circuit is the selection of the output of two receivers, tuned to the same channel, that has the better signal to noise ratio. This circuit compares the two noise leads from the receivers and selects the one with the lower audio noise level.

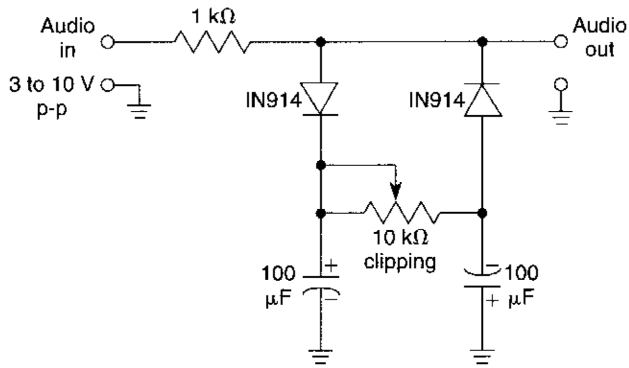


Except as indicated, decimal values of capacitance are in microfarads (μF); resistances are in ohms; k=1,000, M=1,000,000.

IC pins not shown are unused.

Fig. 59-3

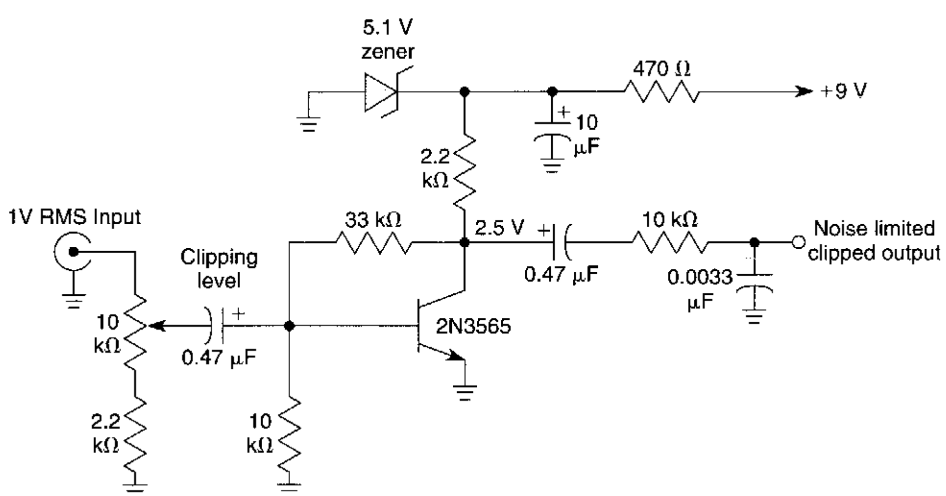
ADJUSTABLE NOISE CLIPPER



WILLIAM SHEETS **_Fig. 59-4_**

This circuit uses two diodes and is a peak-to-peak limiter. The capacitors charge up to a dc level determined by the peak-to-peak audio signal and the clipping control. A positive or negative peak or spike is clipped if it exceeds this level plus the diode drops. The circuit should be operated at several volts level for best results.

SIMPLE NOISE LIMITER



WILLIAM SHEETS **_Fig. 59-5_**

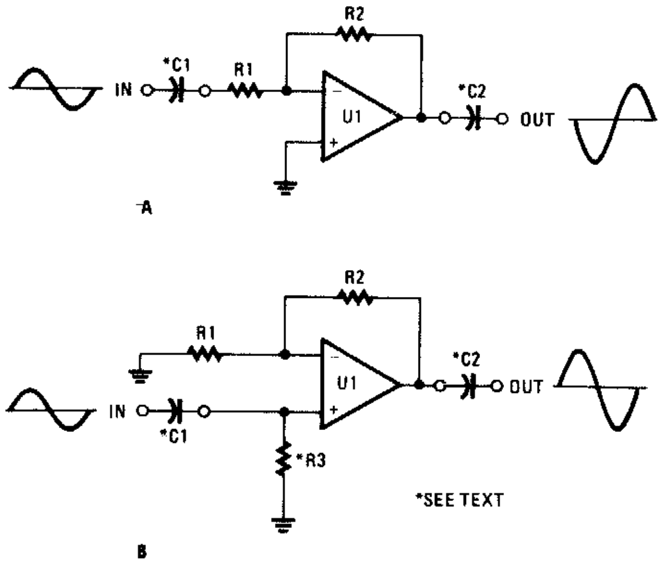
This circuit uses a symmetrical limiter obtained by biasing a transistor to a *Q* point that is half of the supply voltage and driving it into saturation and cutoff. An input of 1 to 2 V RMS is sufficient. This output will be approximately 4 V p-p into a high-impedance load.

60

Operational Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

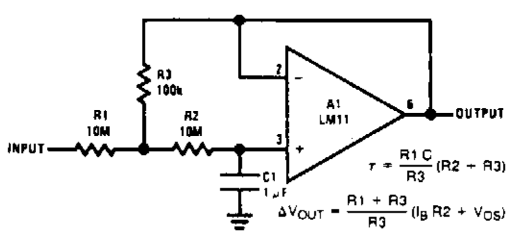
BASIC OP-AMP CIRCUITS



RADIO-ELECTRONICS **_Fig. 60-1_**

The two simplest op-amp configurations are the inverting (A) and the noninverting (B). Resistor R3 is needed only if C1 is used in the noninverting circuit.

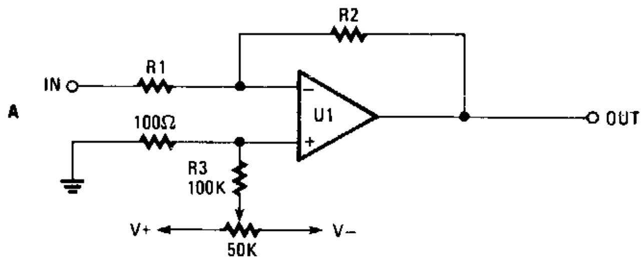
OP AMPS WITH LONG RC TIME CONSTANTS

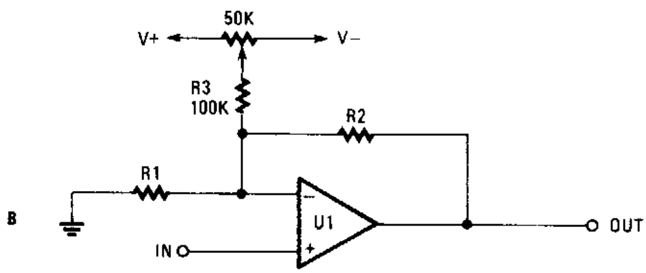


NATIONAL SEMICONDUCTOR **_Fig. 60-2_**

This circuit multiplies RC time constant to 1000 seconds and provides low output impedance. Cost is lowered because of reduced resistor and capacitor values.

OP-AMP OFFSET NULL



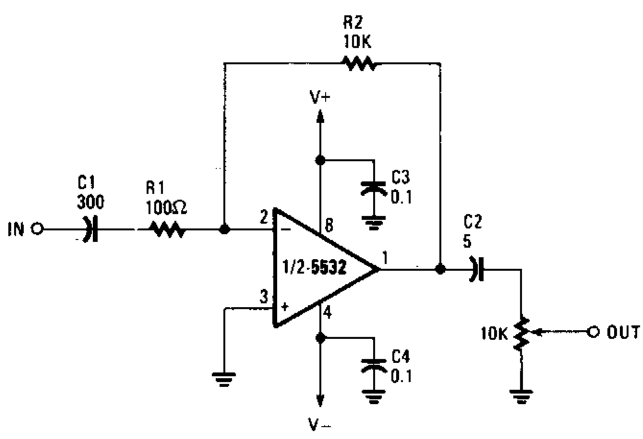


RADIO-ELECTRONICS

Fig. 60-3

Offset problems can occur in the best of circuits (and often do) without regard to whether the circuit is inverting (A), or noninverting (B). Offset-nulling potentiometers are useful in correcting the output to zero, but their effectiveness will vary under different conditions.

BASIC OP-AMP AUDIO AMPLIFIER

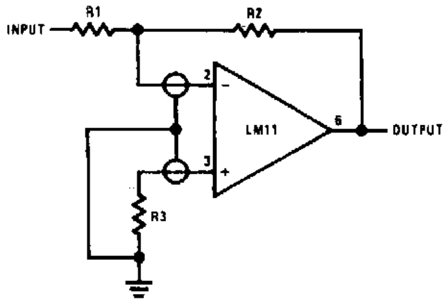


RADIO-ELECTRONICS

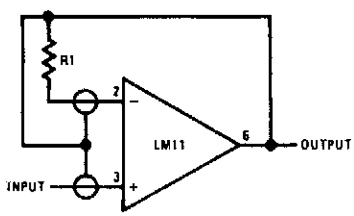
Fig. 60-4

Any general-purpose op amp can be used in this application.

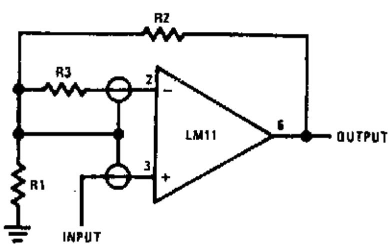
INPUT GUARDING FOR HI-Z OP AMPS



a. Inverting amplifier

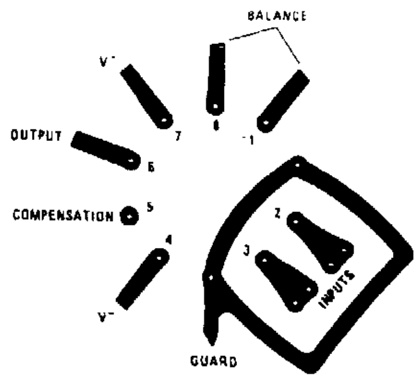


b. follower



c. non-inverting amplifier

Input guarding for various op amp connections. The guard should be connected to a point at the same potential as the inputs with a low enough impedance to absorb board leakage without introducing excessive offset.

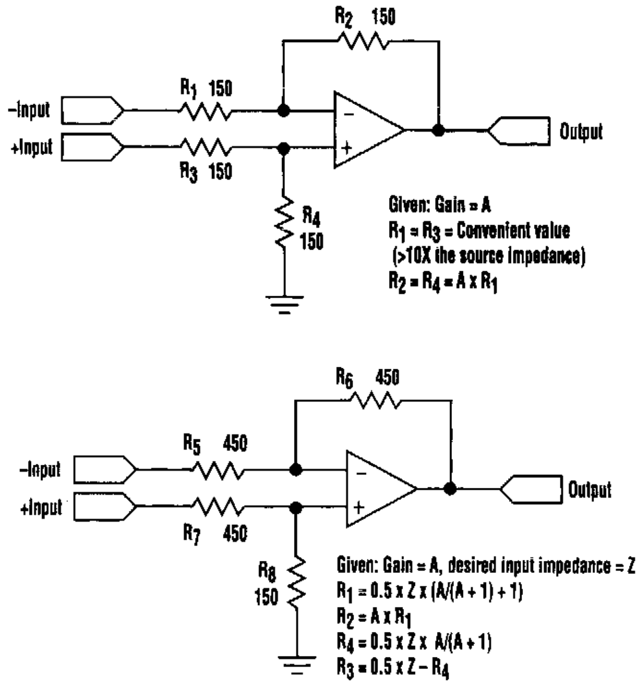


Bottom view

Input guarding can drastically reduce surface leakage. Layout for metal can is shown here. Guarding both sides of board is required. Bulk leakage reduction is less and depends on guard ring width.

NATIONAL SEMICONDUCTOR *Fig. 60-5*

ENHANCED OP-AMP BALANCED AMPLIFIER

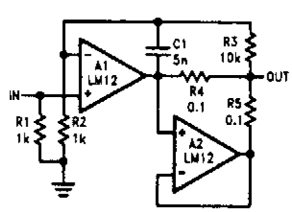


ELECTRONIC DESIGN *Fig. 60-6*

The classic single op-amp balanced circuit works well in low-source-impedance configurations, but tends to struggle in higher-source-impedance applications because of the varying input impedance of the inputs referred to ground.

A modified version of the classic op-amp configuration of the figure uses a different set of formulas to determine the resistor values. It equalizes the impedance of both inputs by considering the op amp's active participation.

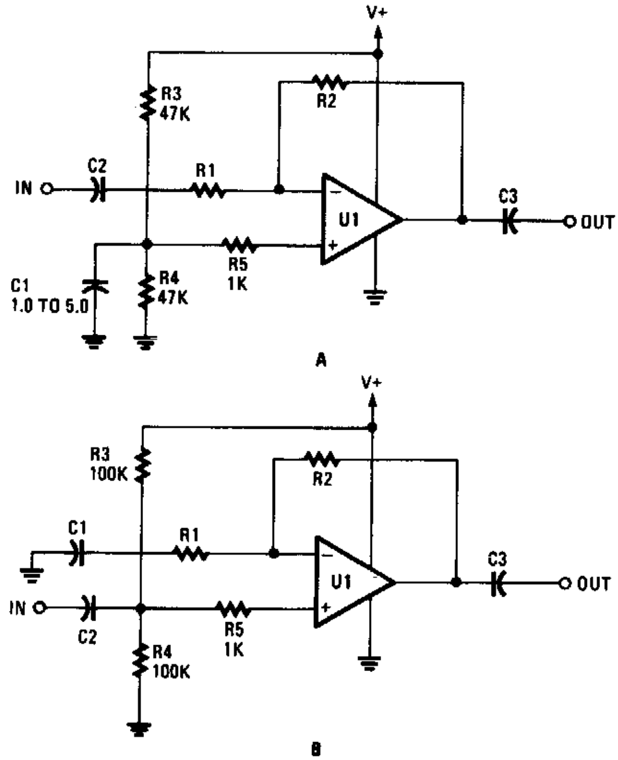
PARALLELED POWER OP AMPS



NATIONAL SEMICONDUCTOR *Fig. 60-7*

Two power op amps can be paralleled using this master/slave arrangement, but high-frequency performance suffers.

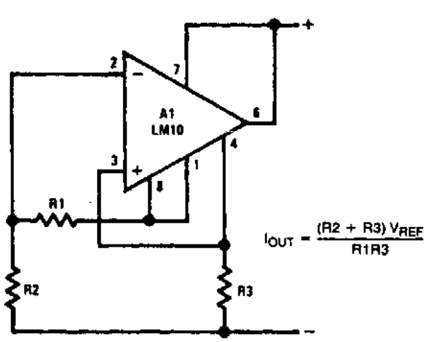
SINGLE-SUPPLY OP-AMP APPLICATIONS



RADIO-ELECTRONICS *Fig. 60-8*

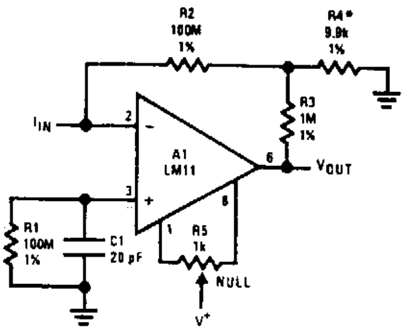
An op amp that normally needs two supplies can be used when only a single supply is needed. The value of *V*+ should be twice the minimum allowable values of the positive and negative voltages normally needed. For example, a 12-V single-supply application would require an op amp capable of ±6-V operation.

CURRENT REGULATOR OP AMP



NATIONAL SEMICONDUCTOR *Fig. 60-9*

OP-AMP RESISTANCE MULTIPLICATION CIRCUIT

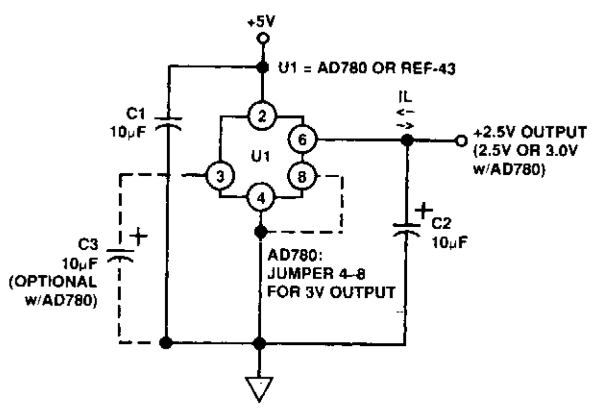


NATIONAL SEMICONDUCTOR

Fig. 60-10

Equivalent feedback resistance is 10 GΩ, but only standard resistors are used. Even though the offset voltage is multiplied by 100, output offset is actually reduced because error is dependent on offset current, rather than bias current. Voltage on summing junction is less than 5 mV.

PSEUDOGROUND



ANALOG DEVICES

Fig. 60-11

For op-amp circuits, a "pseudo ground" is often needed; a voltage reference IC can be used. The Analog Devices AD780 is used here for this application. This can sink or source current.

61

Oscillators (Audio)

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

EASILY TUNED SINE-WAVE OSCILLATOR

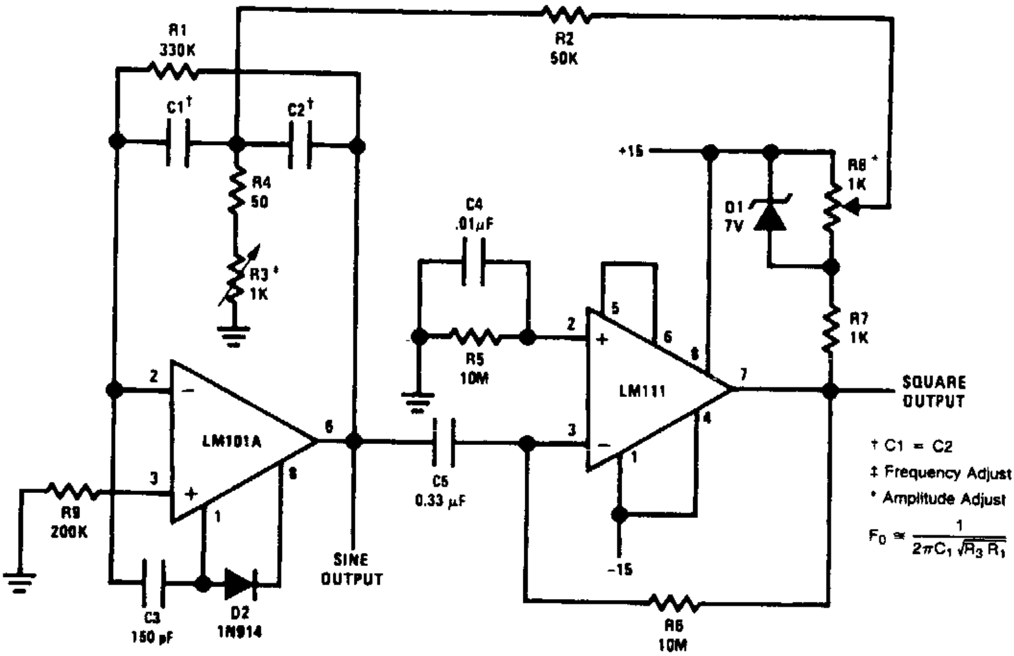


TABLE I

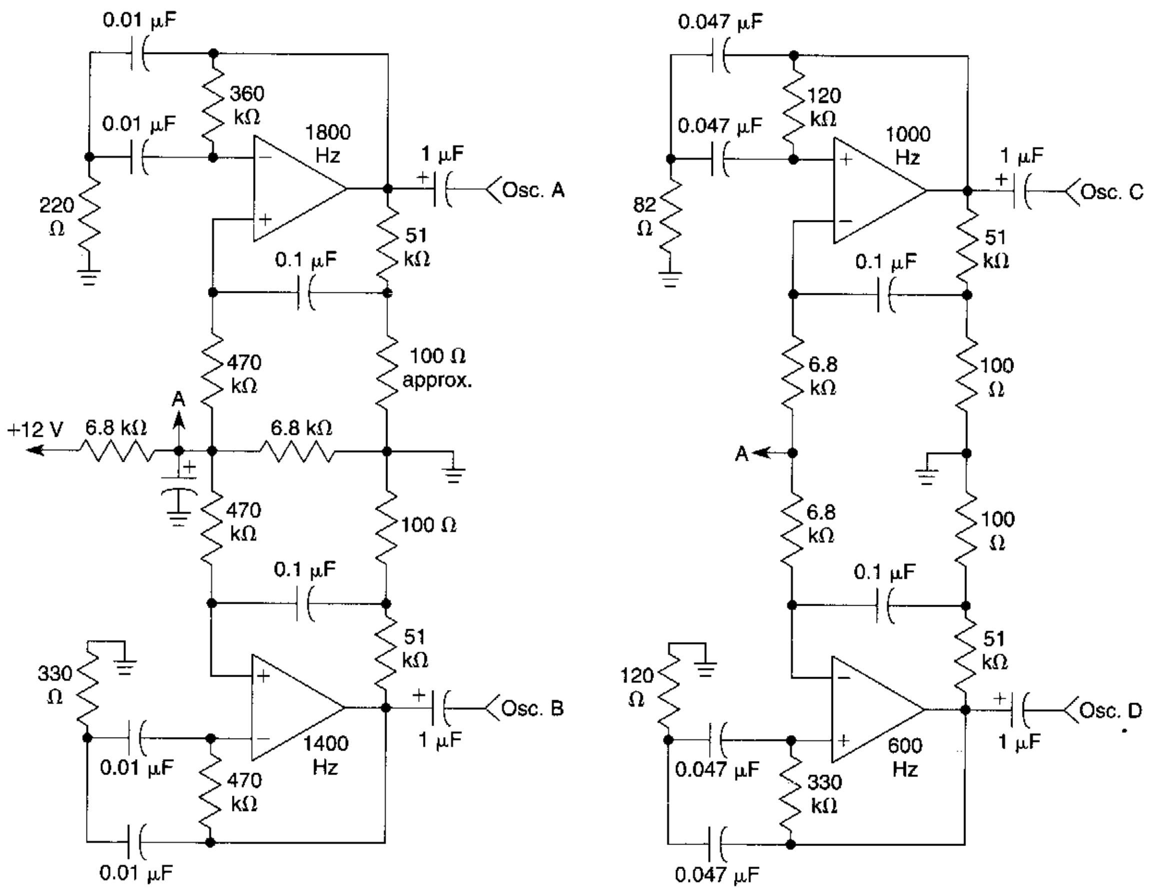
C_1, C_2	Min Frequency	Max Frequency
0.47 μF	18 Hz	80 Hz
0.1 μF	80 Hz	380 Hz
.022 μF	380 Hz	1.7 kHz
.0047 μF	1.7 kHz	8 kHz
.002 μF	4.4 kHz	20 kHz

NATIONAL SEMICONDUCTOR *Fig. 61-1*

The circuit will provide both a sine- and square-wave output for frequencies from below 20 Hz to above 20 kHz. The frequency of oscillation is easily tuned by varying a single resistor. This is a considerable advantage over Wien bridge circuits, where two elements must be tuned simultaneously to change frequency. Also, the output amplitude is relatively stable when the frequency is changed.

An operational amplifier is used as a tuned circuit, driven by square wave from a voltage comparator. Frequency is controlled by R1, R2, C1, C2, and R3, with R3 used for tuning. Tuning the filter does not affect its gain or bandwidth, so the output amplitude does not change with frequency. A comparator is fed with the sine-wave output to obtain a square wave. The square wave is then fed back to the input of the tuned circuit to cause oscillation. Zener diode, D1, stabilizes the amplitude of the square wave fed back to the filter input. Starting is ensured by R6 and C5, which provide dc negative feedback around the comparator. This keeps the comparator in the active region.

QUAD TONE OSCILLATOR

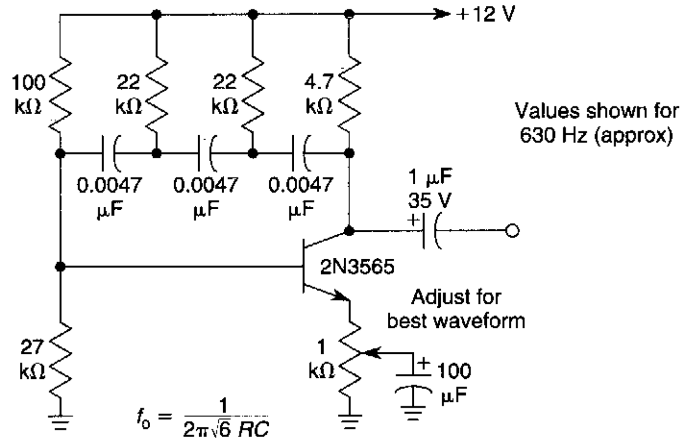


WILLIAM SHEETS

Fig. 61-2

A quad op amp (TL084, etc.) can be used to produce four audio tone generators for use in a test setup. The circuit uses a 12-V supply.

ONE-TRANSISTOR PHASE-SHIFT OSCILLATOR

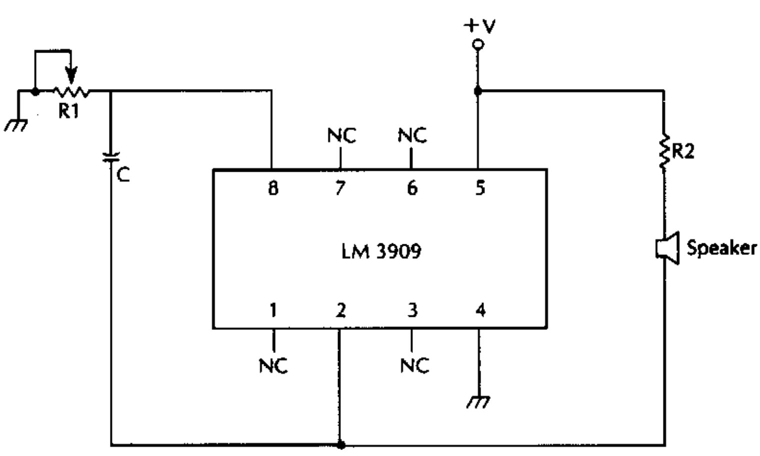


WILLIAM SHEETS

Fig. 61-3

A single transistor is used as an active element in an RC phase shift oscillator.

BASIC LM3909 AUDIO OSCILLATOR

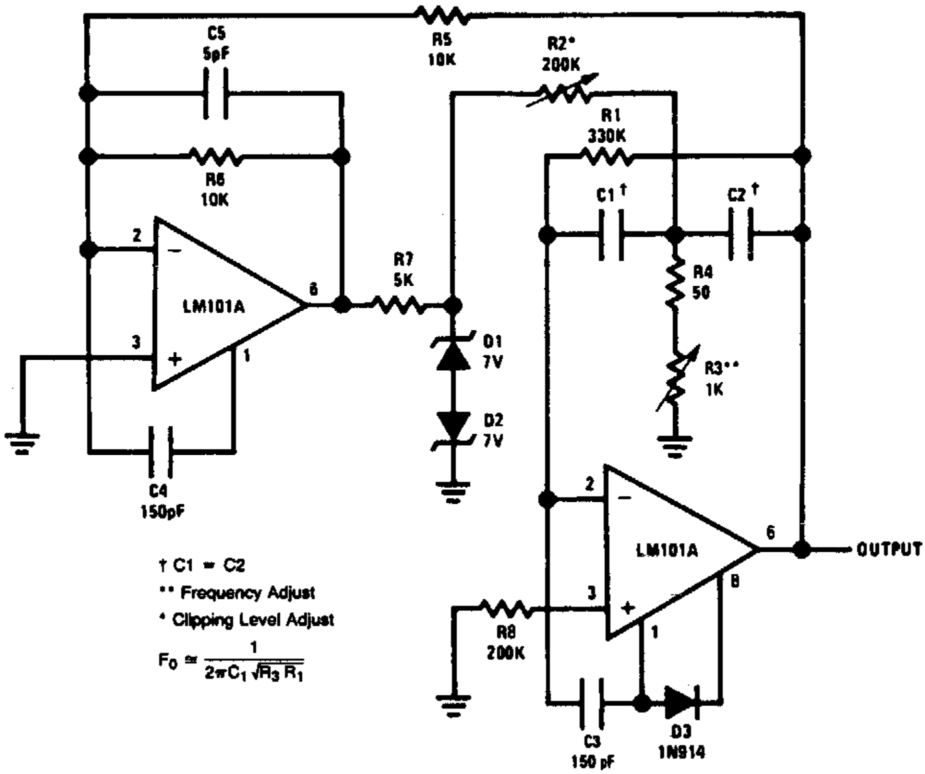


McGRAW-HILL

Fig. 61-4

The LM3909's oscillator frequency can be fine-tuned by adding a resistor to a basic circuit.

LOW-DISTORTION SINE-WAVE OSCILLATOR

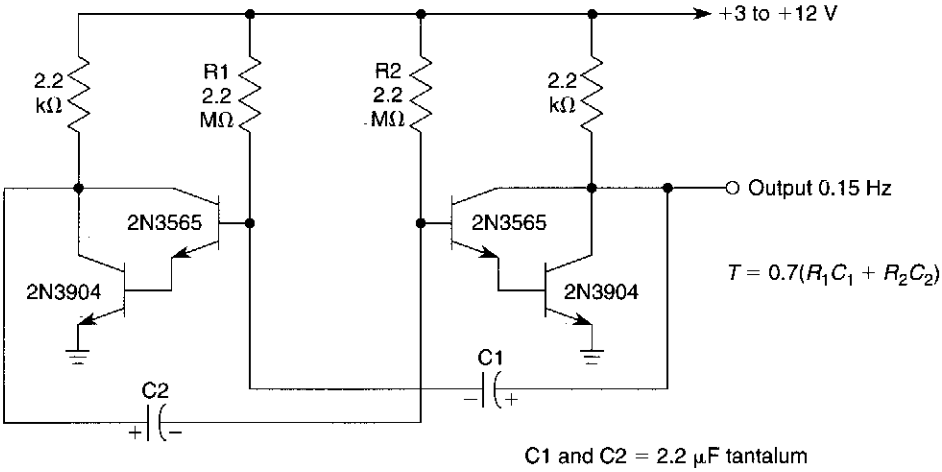


NATIONAL SEMICONDUCTOR

Fig. 61-5

C1, C2	Min. Frequency	Max. Frequency
0.47 µF	18 Hz	80 Hz
0.1 µF	80 Hz	380 Hz
0.022 µF	380 Hz	1.7 kHz
0.0047 µF	1.7 kHz	8 kHz
0.002 µF	4.4 kHz	20 kHz

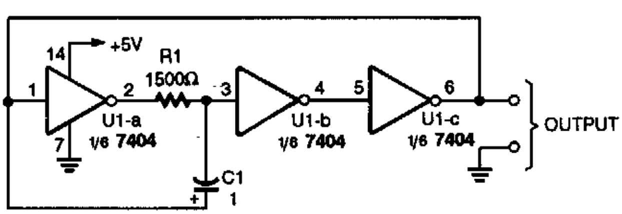
LOW-FREQUENCY ASTABLE



WILLIAM SHEETS *Fig. 61-6*

By using a high-gain low-current transistor, such as the 2N3565, a pair of Darlington-connected transistors (2N3565 and 2N3904) can be used in a high-impedance configuration.

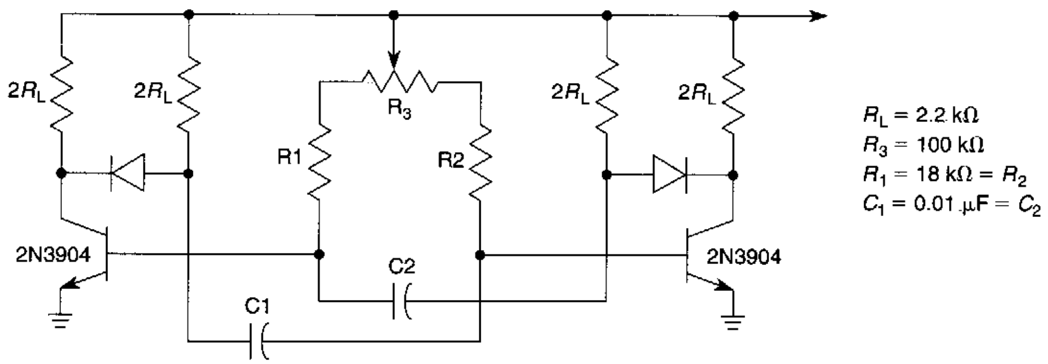
TTL-BASED AUDIO OSCILLATOR



POPULAR ELECTRONICS *Fig. 61-7*

Half a 7404 will produce a tone around 1000 Hz with this circuit.

VARIABLE DUTY CYCLE FROM ASTABLE



WILLIAM SHEETS ***Fig. 61-8***

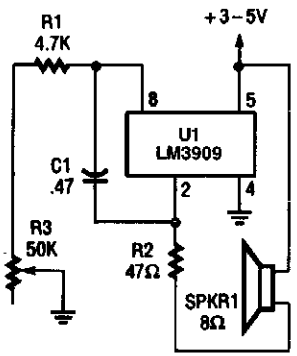
If $R_1 = R_2 = R_3$ and $C_1 = C_2 = C_3$

If potentiometer R3 is set at N% of rotation, then

$T_{\mathrm{TOTAL}} \approx 0.7\ [(R + NR_3)C + [R + (1 - N)R_3C]$

$T_{\mathrm{TOTAL}} \approx 1.4\ (R + R_3)C$ and the duty cycle can be varied without changing frequency.

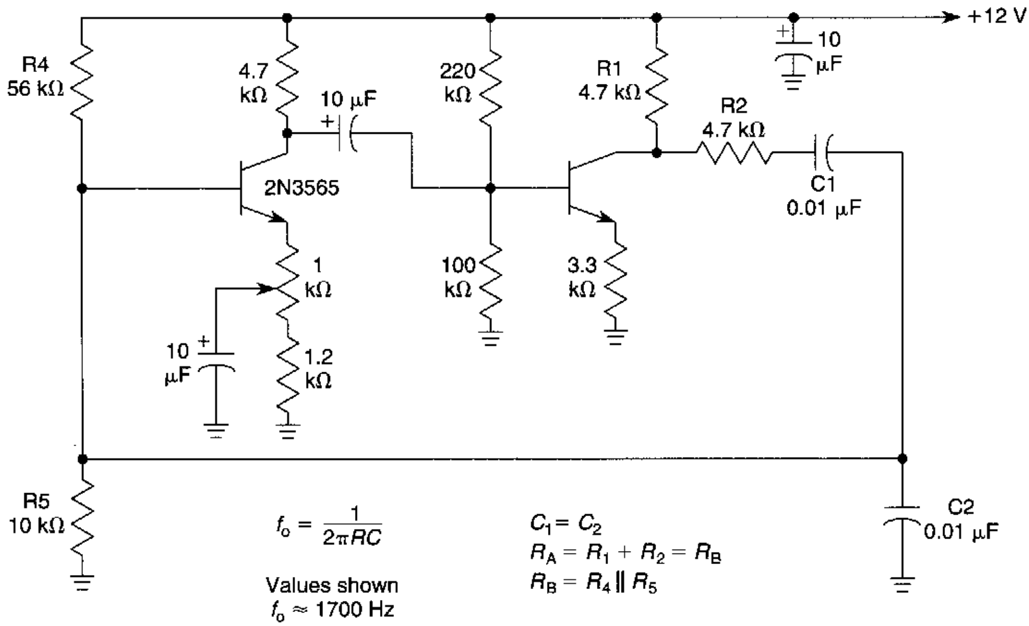
SIMPLE VARIABLE-FREQUENCY OSCILLATOR



POPULAR ELECTRONICS ***Fig. 61-9***

In this variable audio frequency oscillator, the output of U1 at pin 2 is used to drive an 8-Ω speaker through R2 (which functions as a current-limiter).

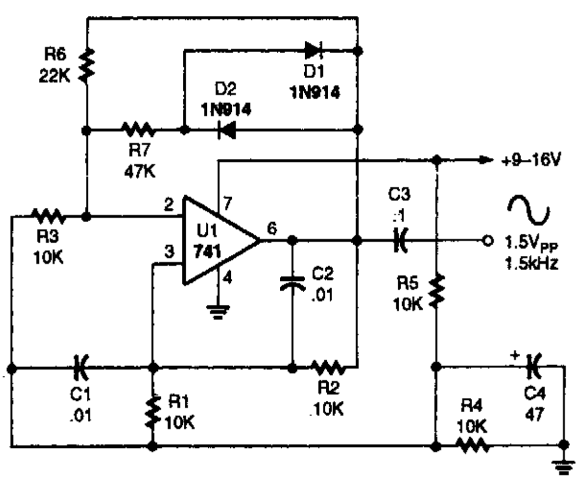
WIEN-BRIDGE OSCILLATOR I



WILLIAM SHEETS

Fig. 61-10

WIEN-BRIDGE OSCILLATOR II

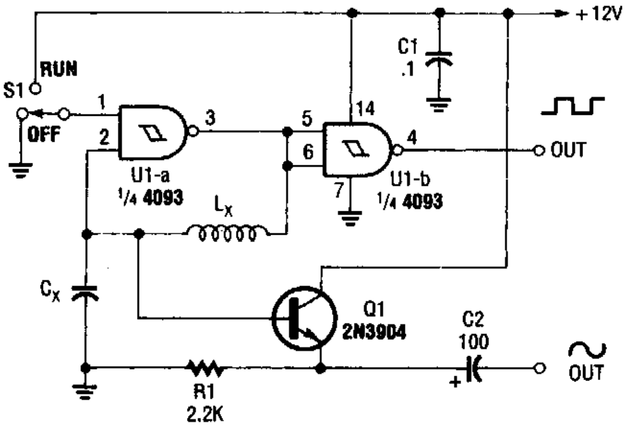


POPULAR ELECTRONICS

Fig. 61-11

The operating frequency of this Wien-bridge oscillator is determined by C1, C2, R1, and R2. It can easily be modified to act as a tunable oscillator by substituting a dual-gang linear potentiometer for R1 and R2.

LOGIC-GATE SINE-WAVE OSCILLATOR



*SEE TEXT

C_x	L_x	f_{OUT}
.018	50mH	9kHz
.018	2mH	14kHz
.047	5mH	5.5kHz
1	1H	300Hz
1	10H	100Hz

POPULAR ELECTRONICS *Fig. 61-12*

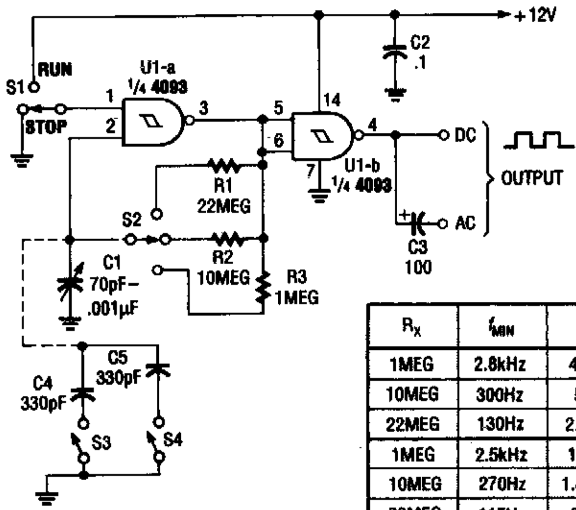
An inductor and capacitor are used here as frequency-determining elements in an LC oscillator.

62

Oscillators (Miscellaneous)

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

VARIABLE-FREQUENCY ASTABLE I

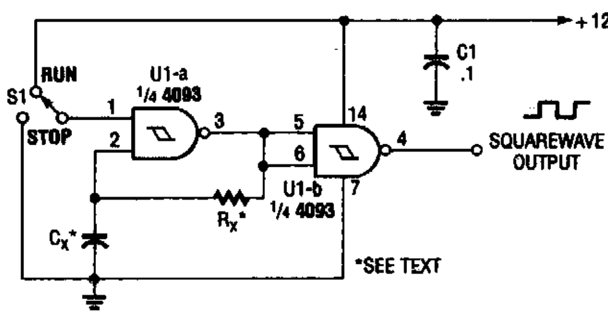


R_X	f_{MIN}	f_{MAX}	
1MEG	2.8kHz	45kHz	C=70pF–.001μF
10MEG	300Hz	5kHz	
22MEG	130Hz	2.5kHz	
1MEG	2.5kHz	17kHz	C=170pF–.0011μF
10MEG	270Hz	1.85kHz	
22MEG	115Hz	835Hz	

POPULAR ELECTRONICS ***Fig. 62-1***

This circuit is a variable-frequency oscillator using a trimmer capacitor or a three-gauge AM broadcast capacitor salvaged from an old AM radio. The three sections must be paralleled.

ASTABLE OSCILLATOR I



POPULAR ELECTRONICS ***Fig. 62-2***

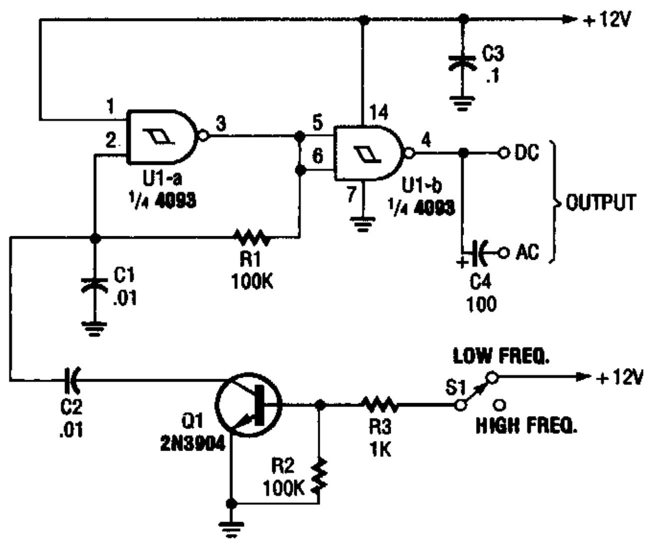
In this circuit, two gates from the quad 4093 package are used to form a simple astable squarewave oscillator.

The values for R_x and C_x are approximately as follows:

C_x	R_x	f_o
0.001 μF	1 MΩ	3 kHz
0.1 μF	1 MΩ	30 Hz
1 μF	10 MΩ	0.03 Hz

These values can be scaled for other frequencies.

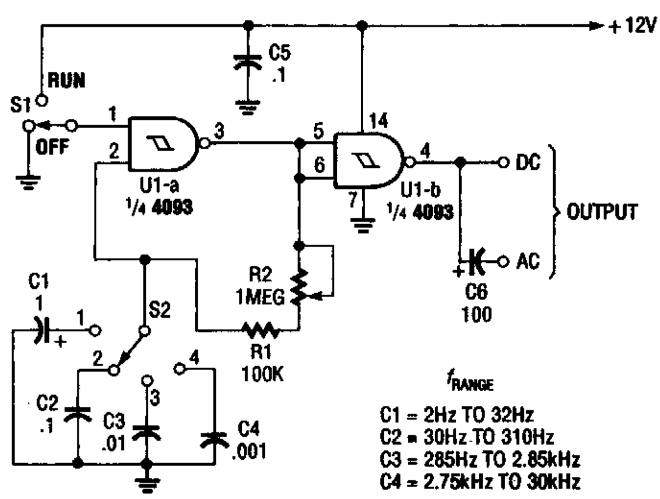
ASTABLE OSCILLATOR II



POPULAR ELECTRONICS *Fig. 62-3*

By using transistor switch Q1/R2/R3, the frequency of an astable oscillator can be changed with a dc voltage or logic level.

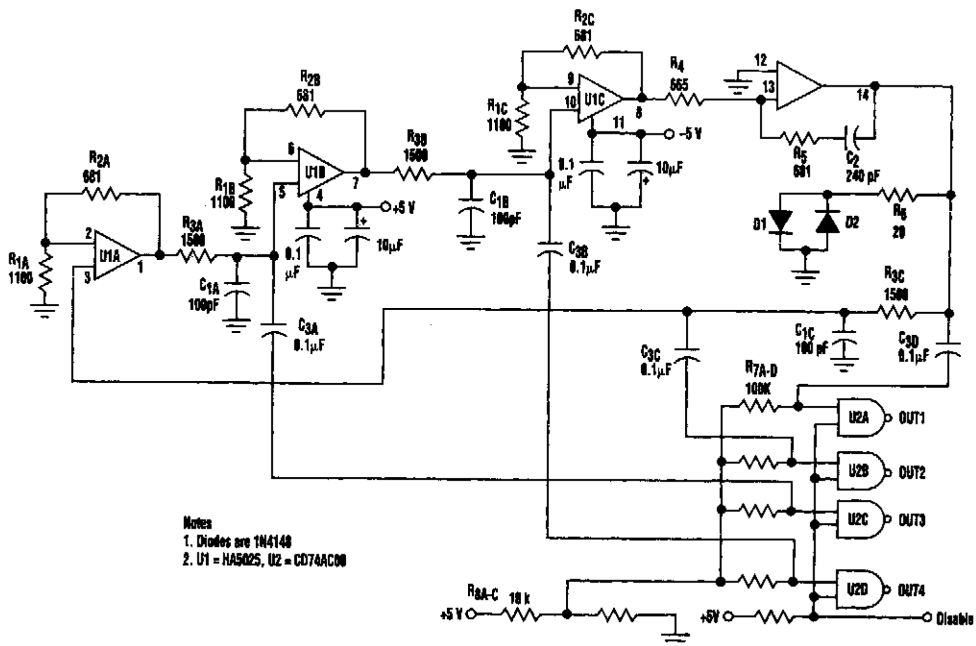
VARIABLE-FREQUENCY ASTABLE II



POPULAR ELECTRONICS *Fig. 62-4*

This circuit uses a single potentiometer and switched capacitors to cover 2 Hz to 30 kHz.

QUADRATURE-WAVE OSCILLATOR

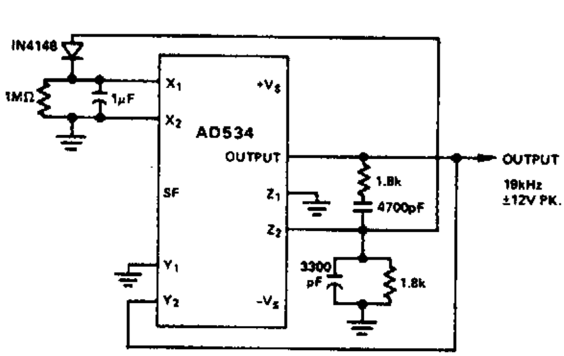


ELECTRONIC DESIGN

Fig. 62-5

By using a high-frequency quad current-feedback amplifier (the HA5025) as an RC oscillator, four quadrature sine waves can be generated. The HA5025's four separate amplifiers generate the sine waves, and the quad NAND gate, U2, is biased at its threshold, so it acts as a sine-wave to square-wave converter when the sine waves are ac coupled into its input.

STABILIZED WIEN-BRIDGE OSCILLATOR

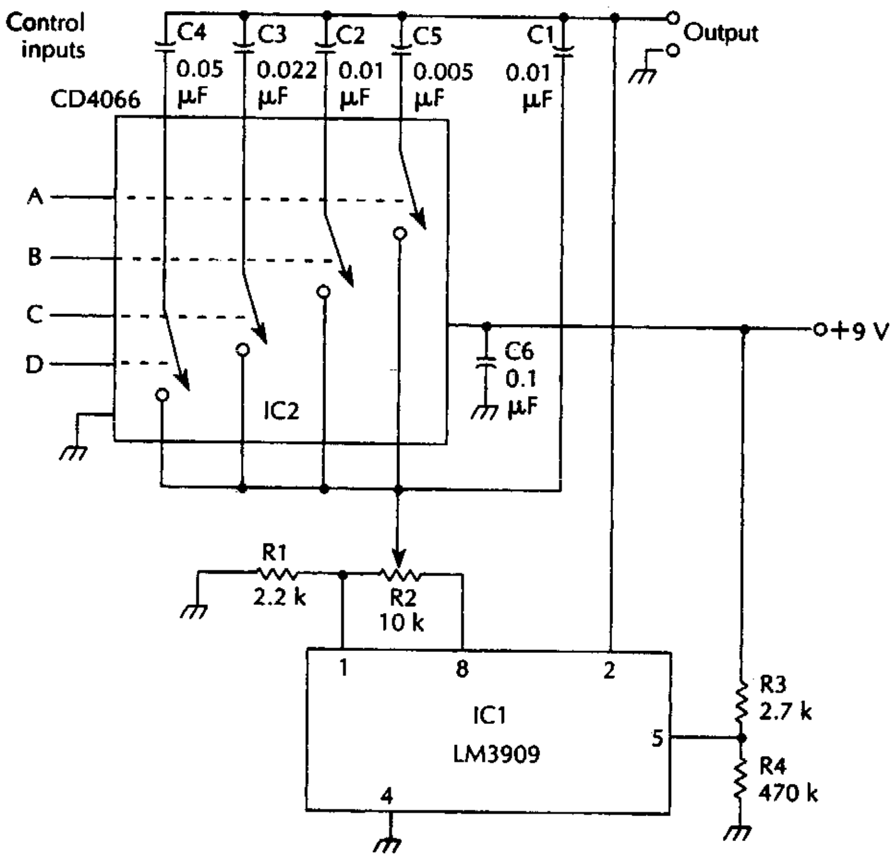


ANALOG DEVICES

Fig. 62-6

In this application, the AD534 is used as a variable-gain amplifier for the feedback signal from the output to the Y input, via the Wien bridge. The peak-rectifier and filter combination applies sufficient voltage to the X (denominator) input to maintain a stable oscillation-amplitude (with about 0.2% ripple). At startup, because X is small (divider mode), the gain is high, and the oscillation builds up rapidly. This is but one of several possible schemes, involving no external active elements. Its forte is simplicity, rather than high performance; nevertheless, the amplitude is not greatly affected by supply and temperature variations, about 0.003 dB per volt, and 0.005 dB per degree.

DIGITALLY CONTROLLED SQUARE-WAVE OSCILLATOR

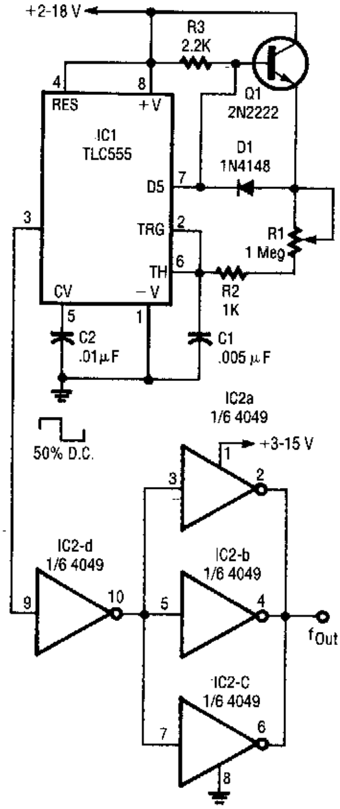


McGRAW-HILL

Fig. 62-7

IC1	LM3909 LED flasher/oscillator	C5, C6	0.1-µF capacitor
IC2	CD4066 quad bilateral switch	R1	2.2-kΩ, ¼-W 5% resistor
C1, C2	0.1-µF capacitor	R2	10-kΩ potentiometer
C3	0.033-µF capacitor	R3	2.7-kΩ, ¼-W 5% resistor
C4	0.047-µF capacitor	R4	470-Ω, ¼-W 5% resistor

50% DUTY-CYCLE 555 CIRCUIT

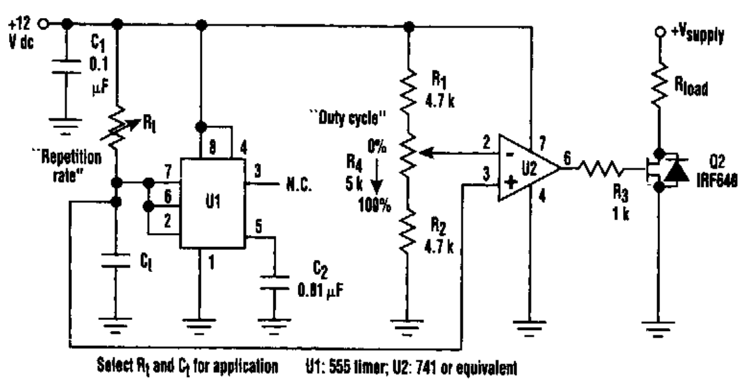


Using an external charge transistor and diode enables a 50% duty cycle and easy frequency control. When the 555's discharge transistor is cut off, the 2N2222 acts as an emitter follower. When the discharge transistor turns on, the 2N2222 turns off and C1 discharges through (R_1 + R_2) at the same rate. The IN4148 provides temperature compensation.

ELECTRONICS NOW

Fig. 62-8

VARIED REP RATE, DUTY CYCLE WITH 555

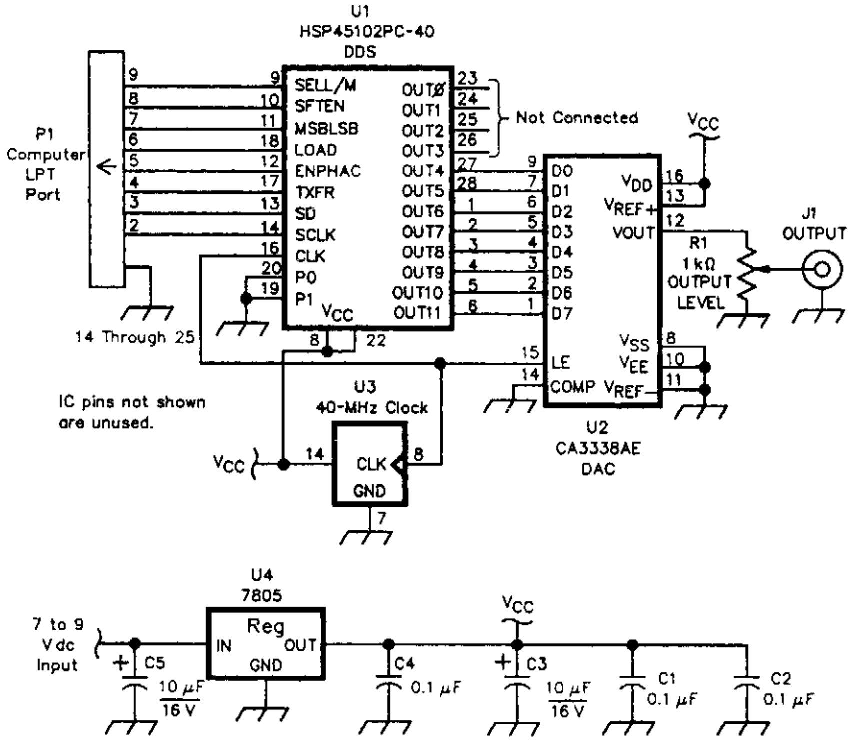


To independently vary rpm and dwell, or duty cycle, a 555 timer is used to produce a ramp waveform, which is compared to an adjustable reference.

ELECTRONIC DESIGN

Fig. 62-9

DDS DIGITAL VFO



QST *Fig. 62-10*

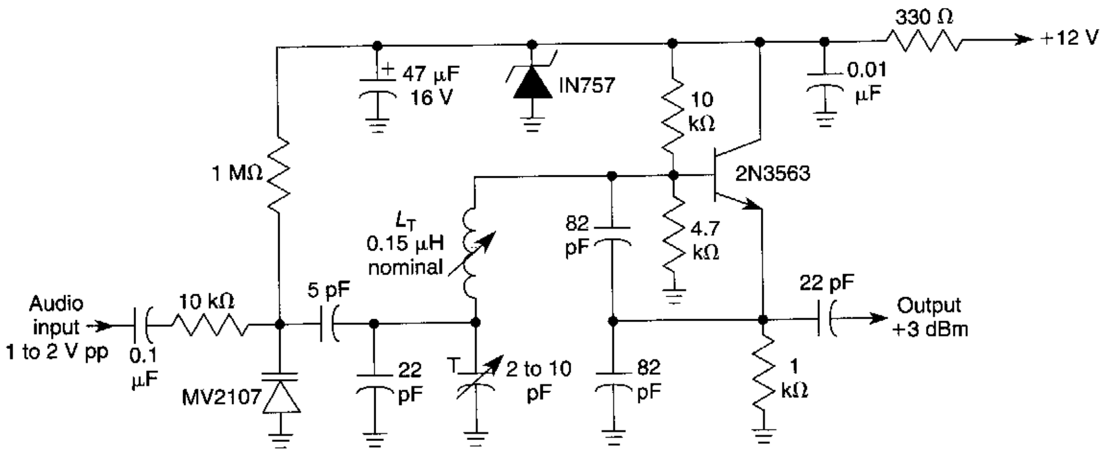
The DDS chip (U1) generates a data stream that is converted by D/A converter U2 into a sine wave. U1 is programmed via the input from P1, from the LPT port of an IBM PC. The system uses a 40-MHz TTL output clock module.

63

Oscillators (RF)

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

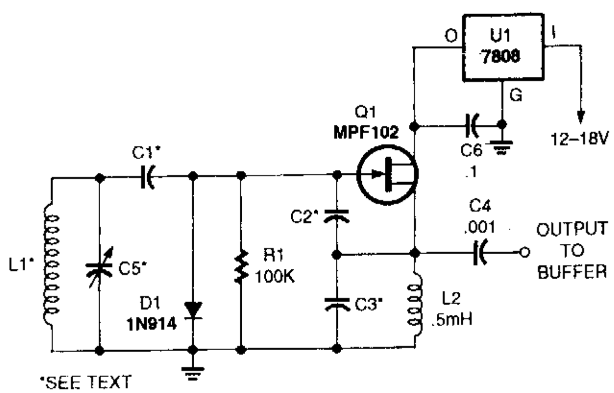
FREQUENCY MODULATED OSCILLATOR



WILLIAM SHEETS *Fig. 63-1*

This circuit can be used for FM wireless audio, microphone, and part-15 applications where a stable frequency modulated oscillator is needed..LT can be varied to cover 75 to 150 MHz, as needed.

JFET VARIABLE-FREQUENCY OSCILLATOR

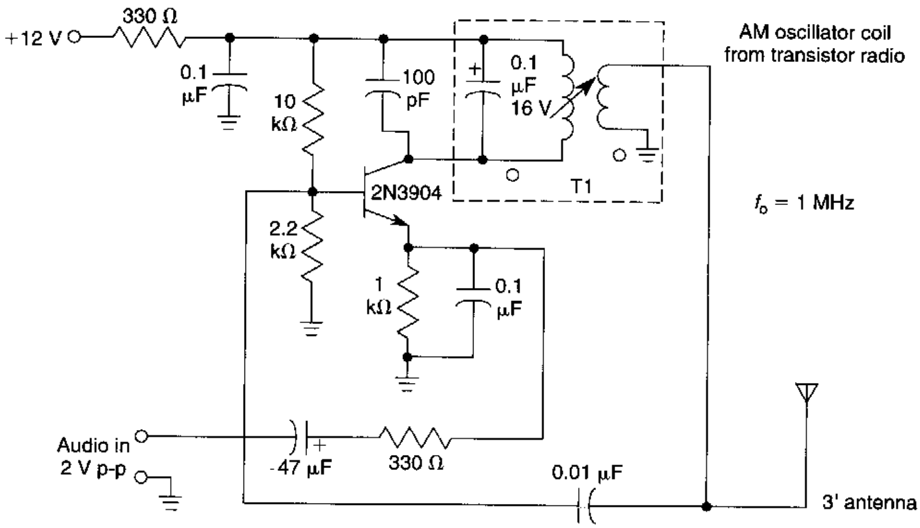


f_OMHZ	L1 (TURNS)	C2 AND C3 (pF)	C1 (pF)	VARIABLE C5(pF)
2–6	36	680	150	100
5–8	25	470	100	100
7–14	20	470	100	50

POPULAR ELECTRONICS *Fig. 63-2*

This simple JFET-based variable-frequency oscillator can be used in receiver and transmitter circuits.

AM OSCILLATOR FOR WIRELESS MICROPHONES

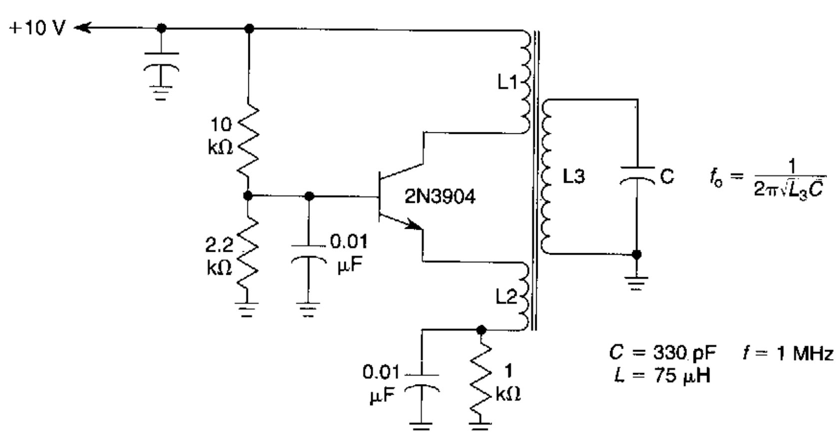


WILLIAM SHEETS

Fig. 63-3

This circuit will generate an AM-modulated signal in the AM broadcast band that can be picked up on a receiver. About 2 V of audio input will produce about 30% modulation of the oscillator signal. An old AM broadcast oscillator coil or other two-winding coil with about a 10:1 turn ratio and about 50 to 150 μH inductance can be used for T1.

REINARTZ OSCILLATOR

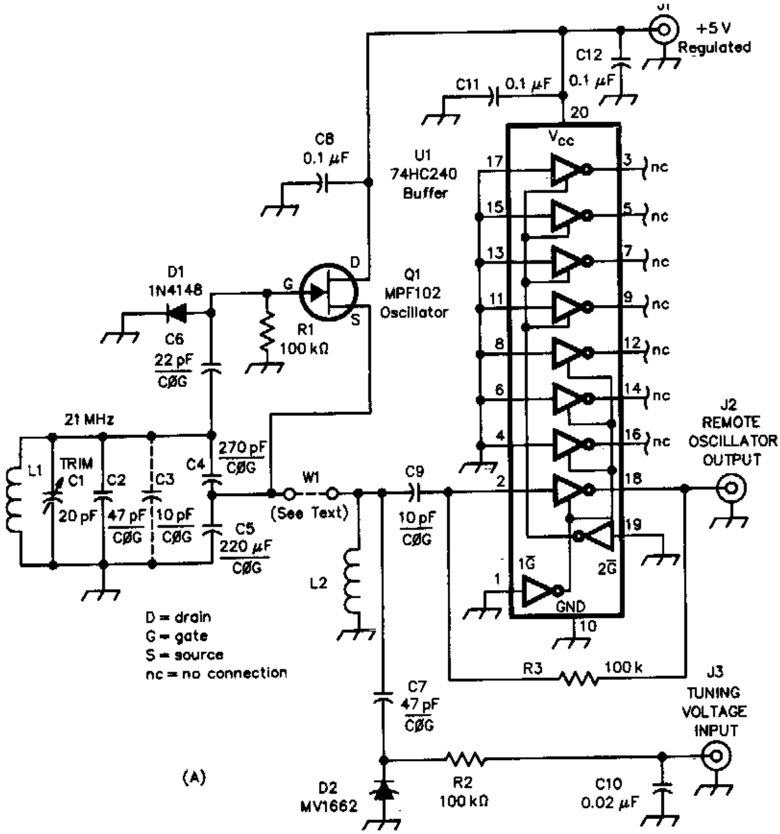


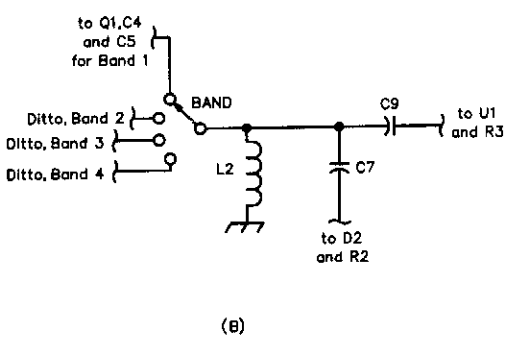
WILLIAM SHEETS

Fig. 63-4

This oscillator uses inductivity coupled emitter and collector windings to its main tank circuit. Take care so that L1 and L2 are not coupled to each other, otherwise this circuit is susceptible to parasitic oscillation at other frequencies. Typically, L1 has 5 to 10 times the number of turns that L2 has. L1, L2, L3 are wound on same coil form. This oscillator is more suited to lower frequencies, ≤10 MHz.

REMOTE-OSCILLATOR HIGH-FREQUENCY VFO





Band-Specific Oscillator Component Values

Band (m)	*L1 (turns*)*	*Approximate Tuning Range† (kHz)*
10	3	110
12	4	100
15	5	80
17	6	65
20	7	50
30	10	40

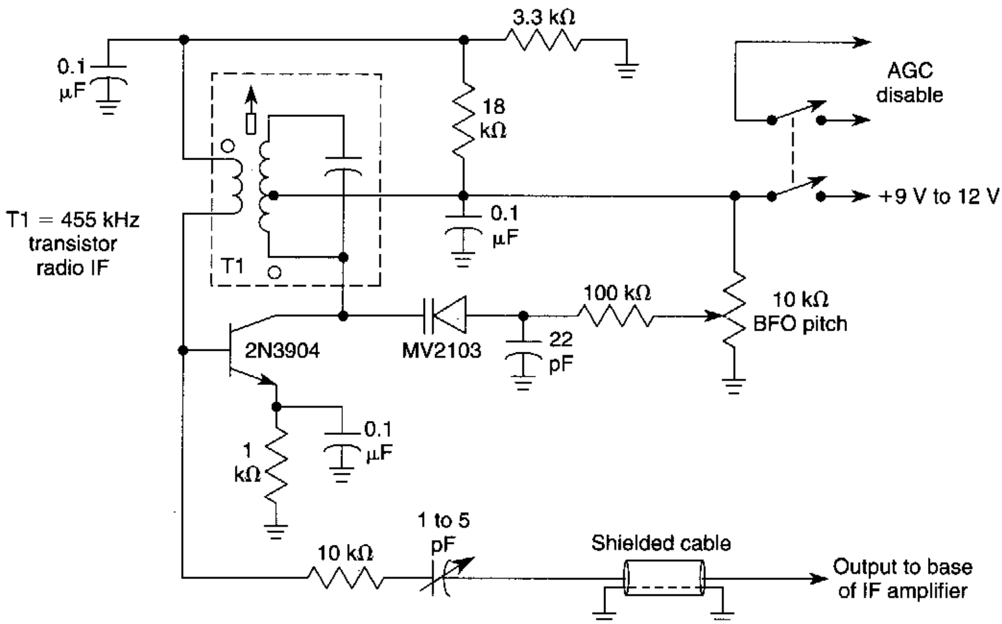
*Close-wound #20 enameled wire on a 3/8-inch plastic rod; see text.

†With an MV1662 diode used at D1, and C7 (Figure 2) equal to 47 pF. See text, Note 2 and the Figure 2 parts list.

QST ***Fig. 63-5***

A remote VFO is sometimes used to control a transmitter or receiver. The circuit shown uses an MPF102 FET and is controlled by a dc voltage at J3. The table shows values for L_1 for various bands from 30 to 10 meters. U1 serves as a buffer amplifier.

BEAT-FREQUENCY OSCILLATOR FOR AM/SW RADIOS

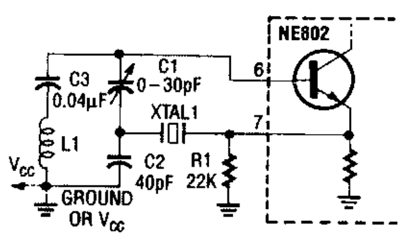


WILLIAM SHEETS

Fig. 63-6

This BFO can be added to inexpensive AM/SW receivers to enable reception of CW signals. Output couples to base of last IF stage. T1 is any 455-kHz IF transformer. The BFO switch should be a DPDT type (as needed), and the radio AGC circuit will probably have to be disabled for CW reception.

BUTLER OSCILLATOR CIRCUIT

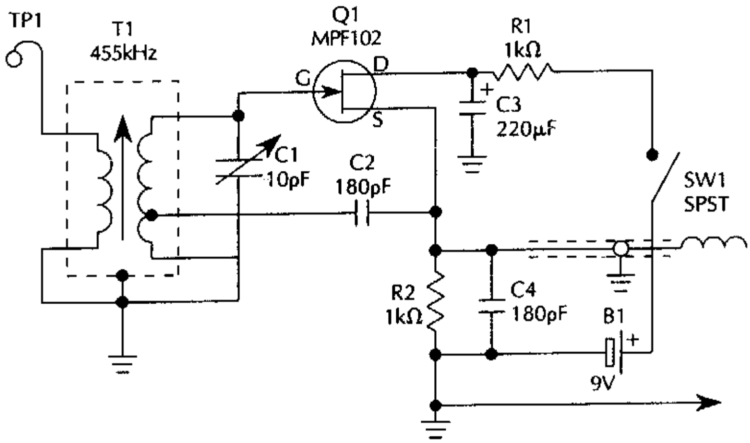


RADIO-ELECTRONICS

Fig. 63-7

This circuit uses an overtone crystal in a Butler oscillator. L1 is approximately 1300 μH, and the crystal frequency should be from 20 to 50 MHz.

455-kHz OSCILLATOR

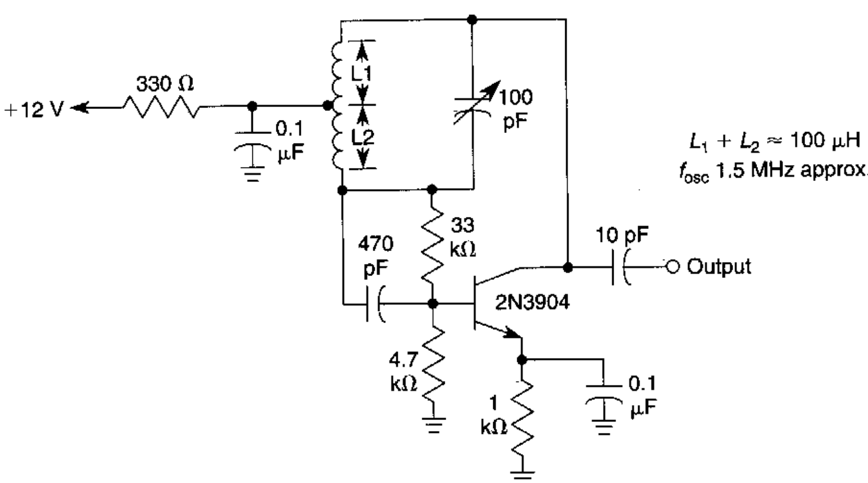


McGRAW-HILL

Fig. 63-8

The 455-kHz oscillator circuit uses a field-effect transistor (FET) for Q1. The output signal is taken from the source circuit of Q1. T1 is a 455-kHz IF transformer.

MODIFIED HARTLEY OSCILLATOR

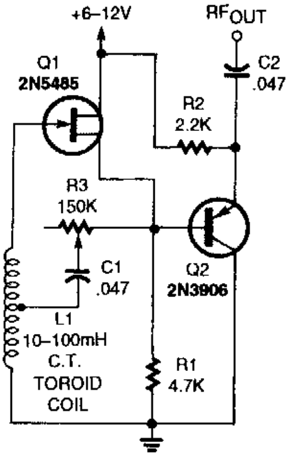


WILLIAM SHEETS

Fig. 63-9

This oscillator uses a tapped coil in the collector circuit, with the tap grounded for the signal. L1 and L2 are coupled inductively and typically have a 3:1 turn ratio, and generally are sections of one entire winding.

VLF LC OSCILLATOR

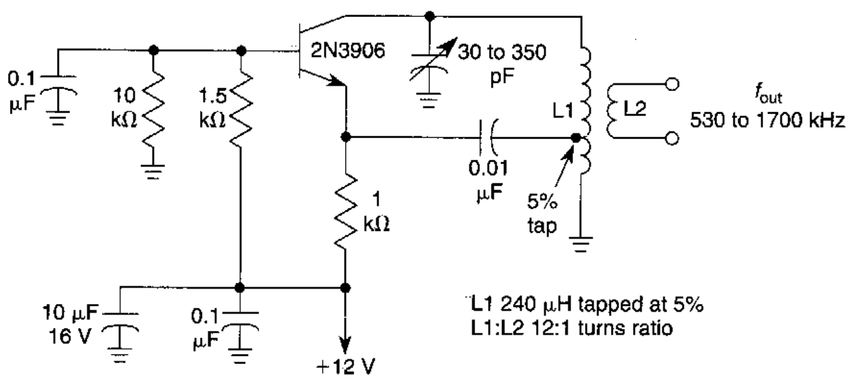


The VLF oscillator uses a large toroid coil as the frequency-determining component and a 2N5485 FET as the active device. R3 is used as a feedback control and also by running the circuit with slightly less feedback than needed for oscillation, can serve as a regenerative amplifier or detector.

POPULAR ELECTRONICS

Fig. 63-10

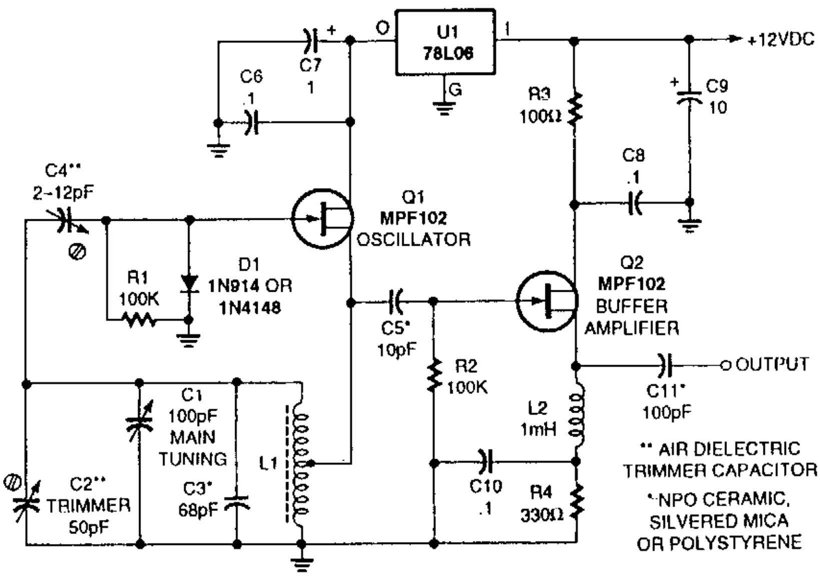
GROUNDED-BASE TUNED COLLECTOR OSCILLATOR FOR AM BROADCAST BAND



WILLIAM SHEETS

Fig. 63-11

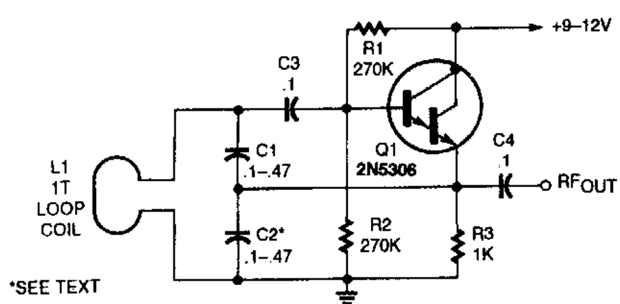
HF VFO CIRCUIT



POPULAR ELECTRONICS *Fig. 63-12*

This typical HF VFO circuit has several stability-enhancing features, including well-chosen capacitor types. The frequency of the VFO is approximately $2\pi (C_1 + C_2 + C_3) L_1$. L1 should be an air-core type coil, rigidly mounted, with high (>200 value) value of Q.

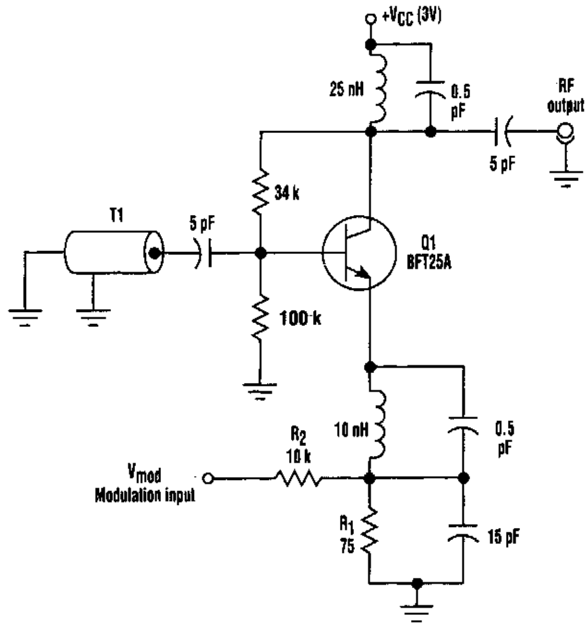
DARLINGTON TRANSISTOR OSCILLATOR



POPULAR ELECTRONICS *Fig. 63-13*

This oscillator uses a very large capacitance-to-inductance ratio. L1 is a one-turn coil consisting of a loop of #12 wire 12" in diameter. This circuit is useful for metal detectors, etc., where a loop antenna is used.

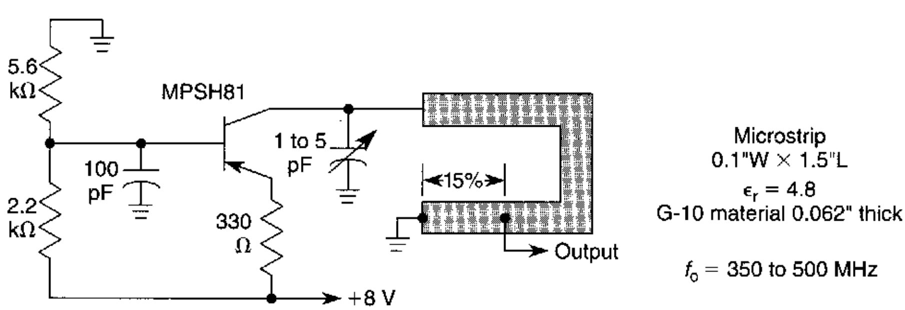
FM HF OSCILLATOR WITH NO VARACTOR



ELECTRONIC DESIGN ***Fig. 63-14***

Instead of using a varactor to frequency-modulate a high-frequency oscillator, this circuit uses base-charging capacitance modulation. Consequently, the large voltage change required by a varactor, which can be a major problem in battery-powered systems with limited supply voltages, is eliminated. T1 is a ceramic coaxial quarter-wave resonator.

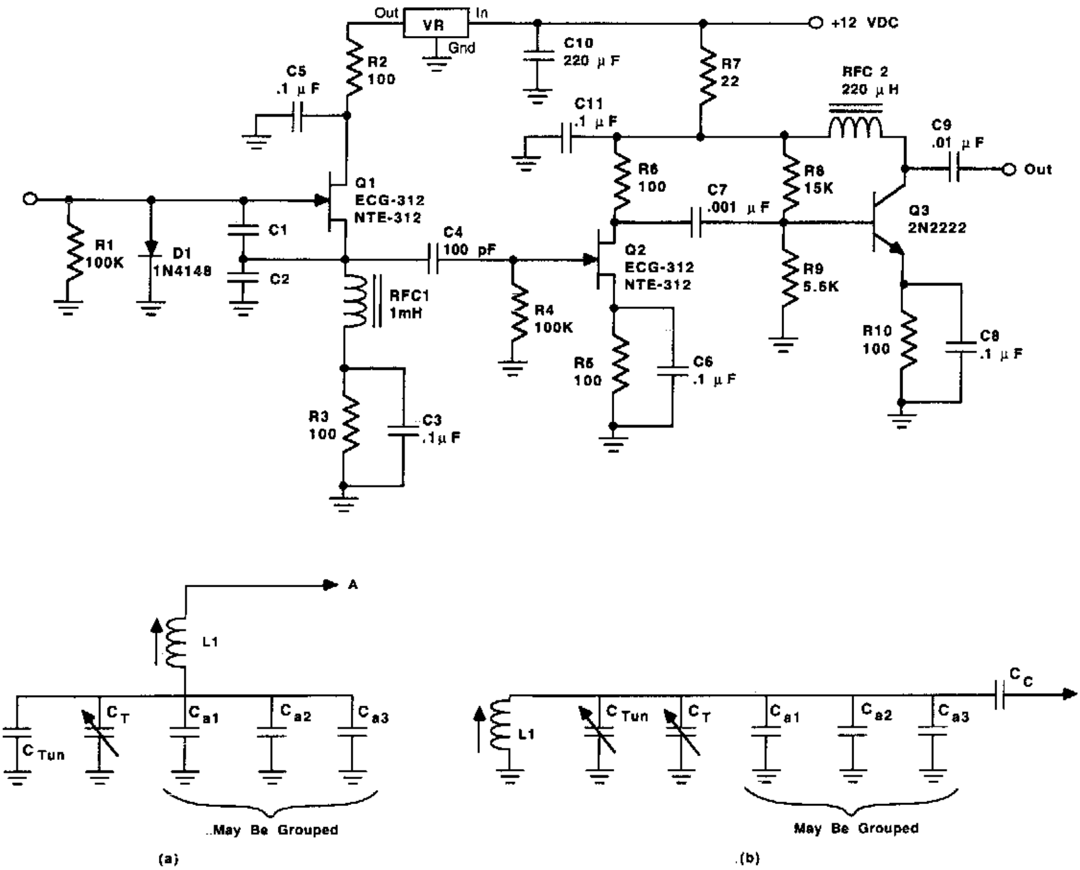
TUNABLE UHF OSCILLATOR



WILLIAM SHEETS ***Fig. 63-15***

This oscillator is typical for 350- to 500-MHz operation. The microstrip inductor is a PC board trace. The tap is typically 15% from the bottom end. The output power is 55 to 100 mW into 50 Ω, with the frequency stability typically 0.1% over 0 to 50°C.

"UNIVERSAL" VFO



73 AMATEUR RADIO TODAY

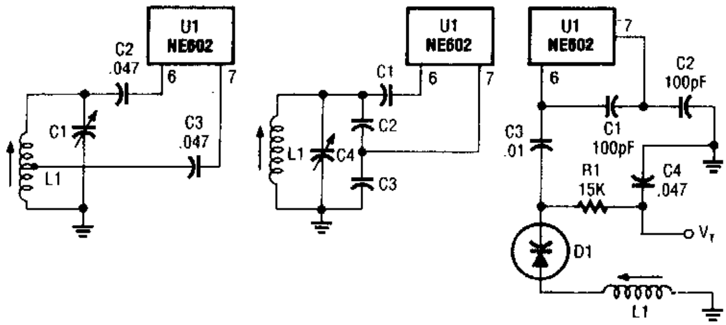
Fig. 63-16

Figure 63-16A shows the basic circuit for the VFO, except for the tuning circuits (which are shown in Fig. 63-16B). Transistor Q1 is a junction field-effect transistor (JFET) oscillator stage. The device to use at Q1 includes MPF-102, 2N4416, and the replacement devices from the popular lines of "service" parts e.g., ECG and NTE).

Two different oscillator configurations can be accommodated by this design (i.e., both Clapp and Colpitts oscillators can be built). Both oscillators are the same from point A in Fig. 63-16C forward, and both depend on a capacitor voltage-divider feedback network. The Clapp oscillator (Fig. 63-16A) is series-tuned and the Colpitts oscillator is parallel-tuned (Fig. 63-16B).

The dc voltage supplied to the oscillator transistor (Q1) is voltage-regulated. The voltage regulator can be any 78Lxx series from 78LO5 to 78LO9.

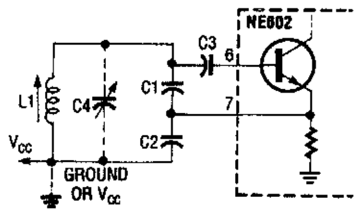
OSCILLATOR CIRCUITS



RADIO CRAFT *Fig. 63-17*

These are methods of using an NE602 with a tunable VFO.

COLPITTS OSCILLATOR

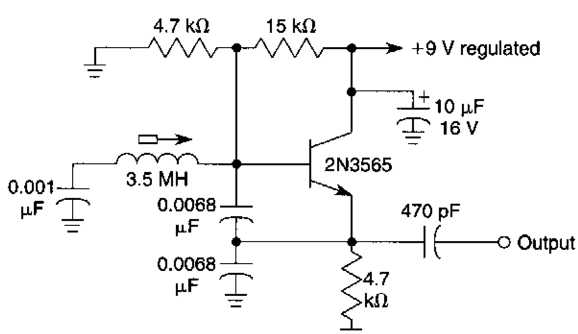


RADIO-ELECTRONICS *Fig. 63-18*

$$L_1 \approx 7\ \mu\text{H}/f(\text{in MHz})$$
$$C_1 \approx C_2 \approx C_3 \approx 2400\ \text{pF}/f$$

In this circuit, the oscillator is free-running.

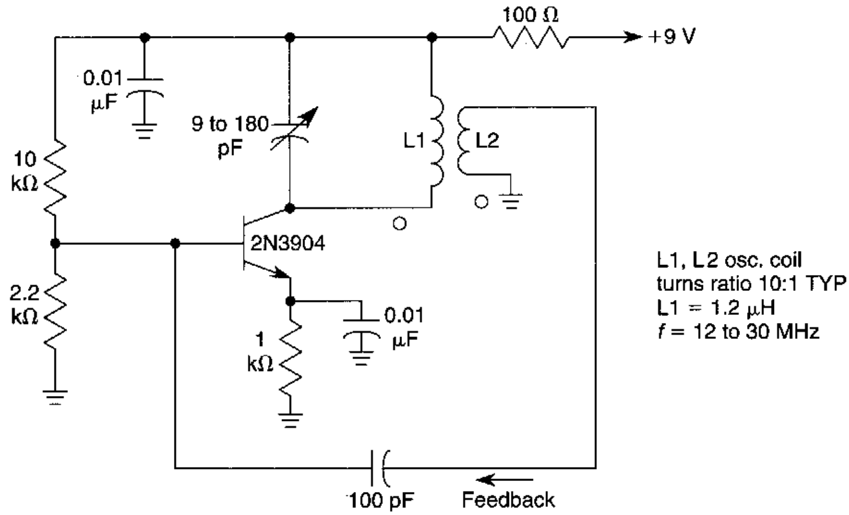
CLAPP OSCILLATOR FOR 100 kHz



WILLIAM SHEETS *Fig. 63-19*

This Colpitts oscillator is very stable and usable where good stability is needed, but crystal control is not desirable. It is capable of 1 part in 104 to 105 with good-quality components.

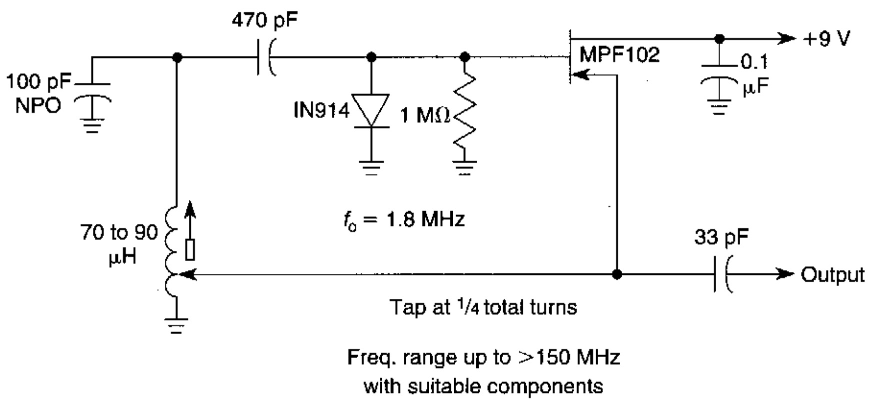
TUNED COLLECTOR OSCILLATOR



WILLIAM SHEETS

Fig. 63-20

HARTLEY OSCILLATOR



WILLIAM SHEETS

Fig. 63-21

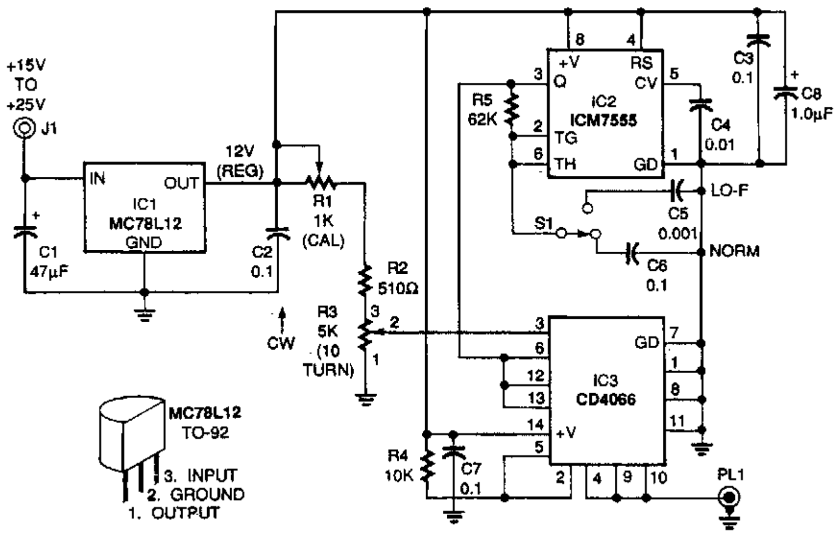
This circuit uses a tapped inductor in a Hartley oscillator circuit. The tap is generally at 25 to 35% total turns in most instances.

64

Oscilloscope Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SCOPE VOLTAGE CURSOR ADAPTER



ELECTRONICS NOW

Fig. 64-1

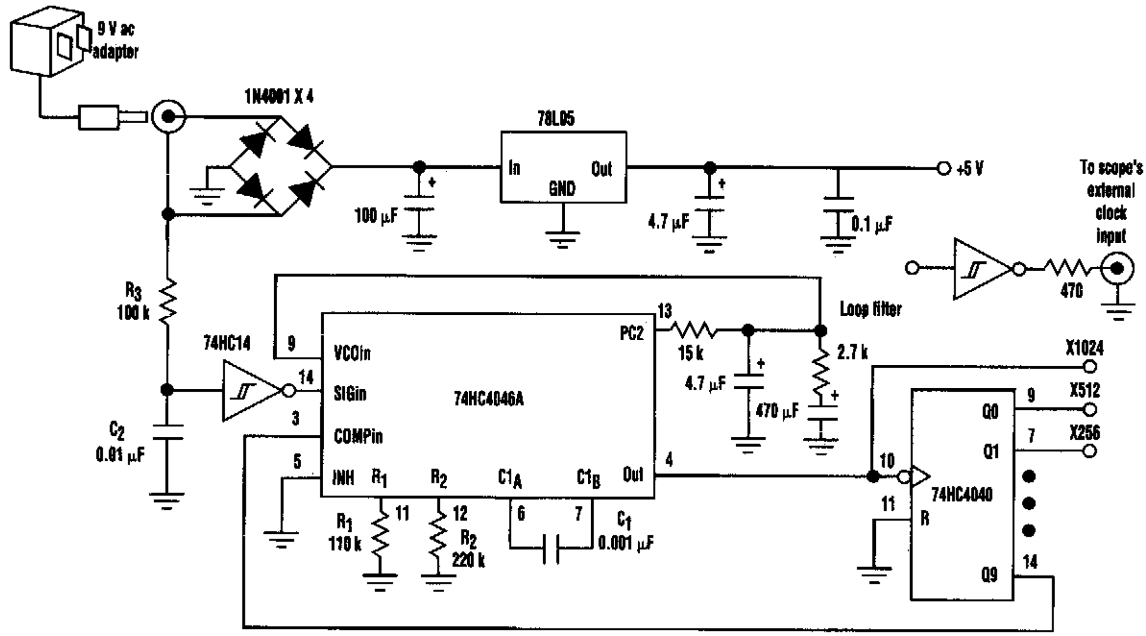
The voltage cursor adapter superimposes horizontal cursor lines on the top and bottom of the waveform—a kind of electronic calipers—to permit direct readout of the voltage value. The cursor lines extend across the entire screen. The MC78L12 voltage regulator (IC1) supplies regulated 12-Vdc to the rest of the circuit. The ICM7555 timer (IC2) drives the CD4066B, a CMOS bilateral switch (IC3). This drive frequency can either be a normal frequency (NORM) of 100 Hz or a low-frequency (LO-*F*) of 10 kHz, depending on the setting of switch S1. Set S1 to LO-*F* for inputs below 500 Hz.

The dc reference voltage supplied to pin 3 of IC3 is set by R3, a 10-turn, 5000-Ω precision potentiometer. The voltage can be read directly from a turns counter dial coupled directly to the potentiometer's wiper. The accuracy of this reading can be 1% or better. Trimmer potentiometer R1 permits the voltage to R3 to be calibrated to precisely 10 V.

The circuit is calibrated by setting the digital reading on the turns counter of R3 to the full clockwise position and adjusting R1 for a reading of 10 V at the wiper of R3 with a digital voltmeter.

Bilateral switch IC3 converts the dc reference to a square wave with exactly the same wiper amplitude. The square-wave output appears on common pins 4, 9, and 10 of IC3 and coaxial plug PL1.

SAMPLING-RATE PHASE LOCK



ELECTRONIC DESIGN *Fig. 64-2*

Most digital scopes have record lengths that are power of 2 (e.g., 1024 points) and sampling rates constrained to a 1-2-5 sequence. This can lead to measurement errors on power-line waveforms because an integral number of line cycles can't be captured. Digital scopes that calculate measurements, such as the rms level, across the entire record will be in error.

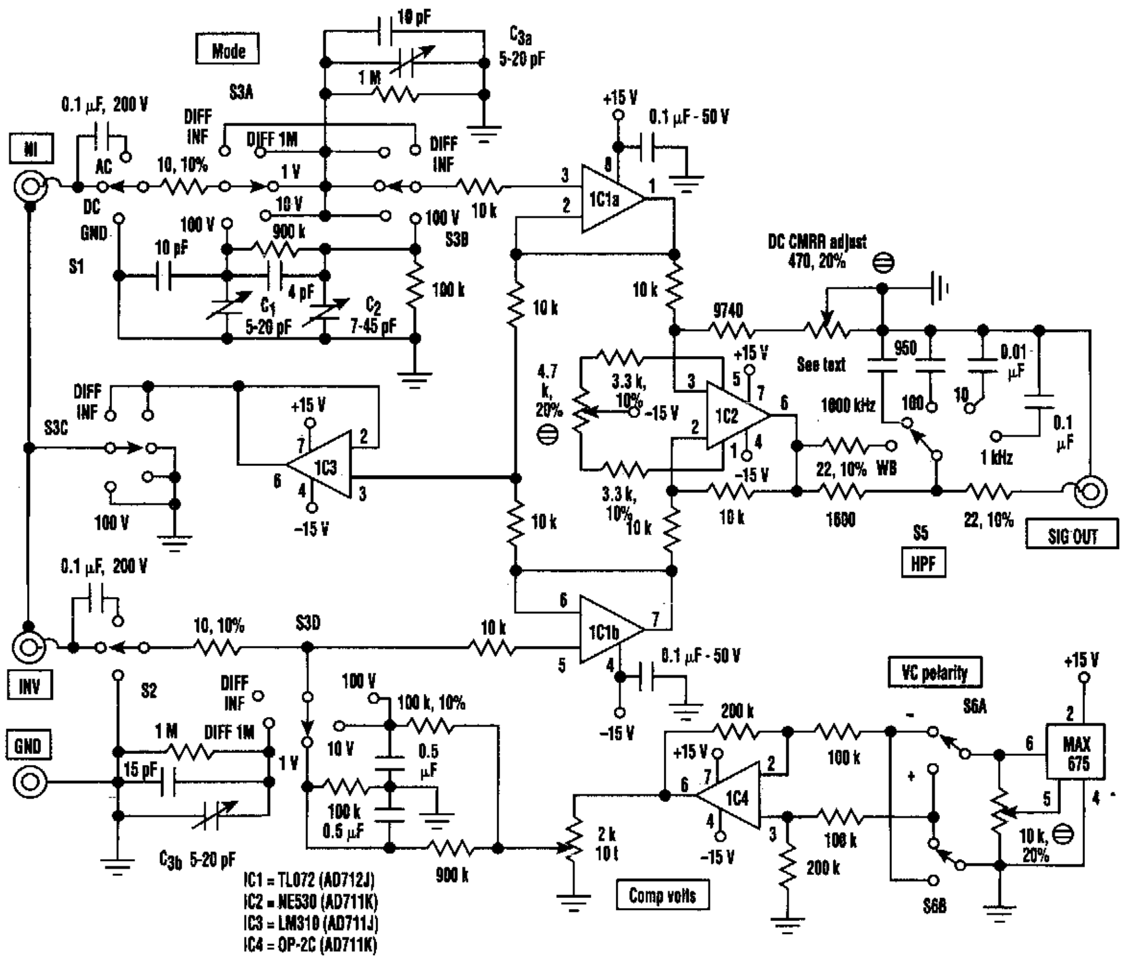
One solution to this problem is to phase-lock the scope's sampling rate to the line frequency by exploiting the external clock input found on some digital scopes. Phase-locking the sampling to line frequency also tracks variations in the power-line frequency.

A 9- or 12-Vac wall transformer provides the circuit's power and the frequency reference. The negative output of the diode bridge refines the circuit ground. The 78L05 regulator provides the +5-V supply for the three ICs. R3 and C2 create a low-pass filter on the half-cycles from one of the floating transformer outputs. R3 also limits the current into the internal diode clamps of the inverter gate. The inverter output becomes the power-line frequency reference and is one input (SIG in to the phase comparator) of the Signetics 74HC4046A phase-locked loop (PLL). The 74HC4040 divides the PLL output frequency by 1024 and feeds the divided clock back to the other PLL phase-comparator input (COMP in). The phase-comparator output (PC2) is filtered and drives the PLL's control voltage (VCO in) so that the output frequency is 1024 times the reference frequency.

With the loop filter shown, the output frequency locks to the line frequency in about 10 s. The oscillator is locked to both 50- and 60-Hz inputs using a 74HC4046A and the values shown for resistors R1 and R2 and capacitor C1.

The output signal is buffered and sent to the scope's external clock input, which is typically a TTL-compatible input. A different tap from the 74HC4040 can be selected to control the number of cycles captured in one scope record.

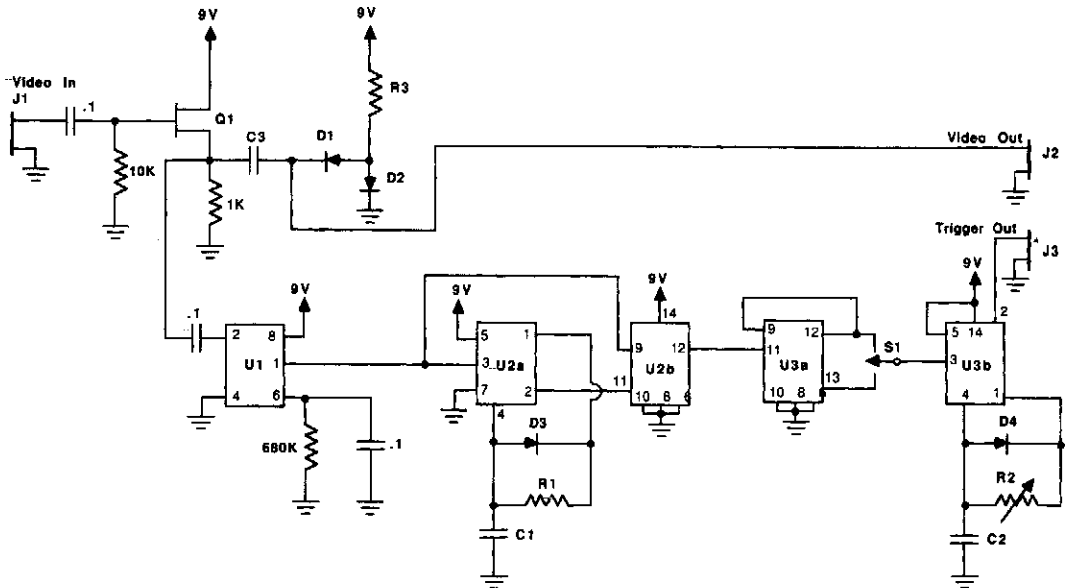
DIFFERENTIAL AMPLIFIER FOR SCOPES



ELECTRONIC DESIGN *Fig. 64-3*

Differential inputs and comparator modes can be added to any general-purpose oscilloscope using this circuit setup. Calibration doesn't change because the circuit operates in unity gain in most modes. Amplifier noise level is low enough not to degrade low-level signals, and its dynamic range can handle signals up to ±12 V peak. Notice that all of the resistors are 1%, unless specified otherwise.

DELAYED VIDEO TRIGGER FOR SCOPES



R1	3.3k	D1,D2,D3,D4	1N914 silicon diode
R2	1 meg potentiometer, linear taper	Q1	MPF102 JFET
R3	1k	U1	LM1881 video sync separator
C1	0.0047 μF	U2,U3	4013 D type flip-flop
C2	0.033 μF	J1,J2,J3	RCA phono jack
C3	0.1 μF	S1	SPDT switch

Note: Resistors are all 5% 1/4 watt. Capacitors are all polyester type.

73 AMATEUR RADIO TODAY *Fig. 64-4*

This circuit will extract vertical sync from a video signal, produce a vertical sync pulse, and add an adjustable delay. This permits a delayed sweep effect to enable a scope to look at any particular horizontal line. It is useful for older scopes.

65

Photography-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

CHARGER FOR PHOTOFLASH CAPACITOR

C1—0.2 μF ±20%, 100 V
C (Load Capacitor)—480 μF, 500 V
D1, D2—MR814 (Fast-Recovery Rectifier)
Q1—MPS6520 (Selected)
Q2—MPS6563 (Selected)
Q3—MPS6562 (Selected)
Q4—MP3613 (Selected)
VR—Neon Lamp (Selected 5 AG)
R1—39K
R2—100Ω
R3—1.0K
R4—120Ω
R5—150Ω
R6—270Ω ±5%
R7—7.5Ω ±5%
R8—1.0 MΩ
R9—2.0 MΩ Pot
R10—390K ±5%
Note: All resistors ±10%, ¼ W, Unless Otherwise Specified

L1: Timing Inductor

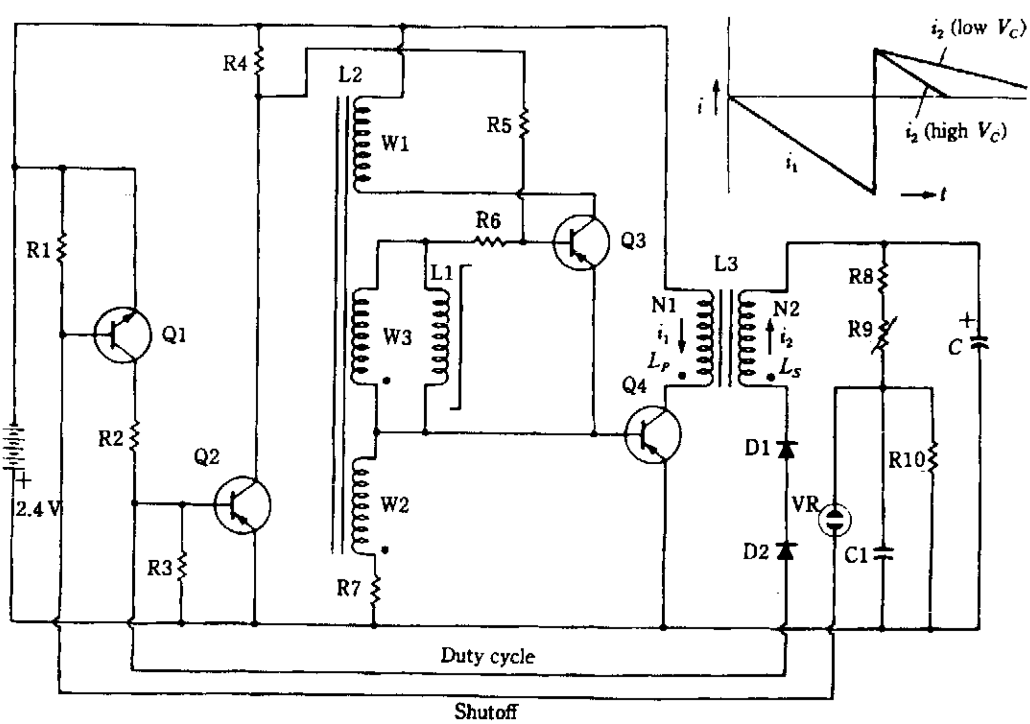
Core:	Ferroxcube 266T125-3E2A
Winding:	145 Turns, No. 36 Wire

L2: Drive-Oscillator Transformer

Core:	Ferroxcube No. 18/11PL00-3B7
Bobbin:	1811F2D
Air Gap:	0.005 in
Windings:	W1: 40 Turns, No. 28 Wire
	W2: 20 Turns, No. 30 Wire
	W3: 140 Turns, No. 36 Wire

L3: Output Transformer

Core:	Ferroxcube No. 26/16P-L00-3B7
Bobbin:	Ferroxcube No. 26/16F2D
Windings:	N1: 11 Turns, No. 18 Wire
	N2: 1100 Turns, No. 38 Wire
Air Gap:	0.030 in



McGRAW-HILL

Fig. 65-1

This circuit charges photoflash capacitor C (480 μF, 500 V) for photoflash usage.

SLIDE STEPPER

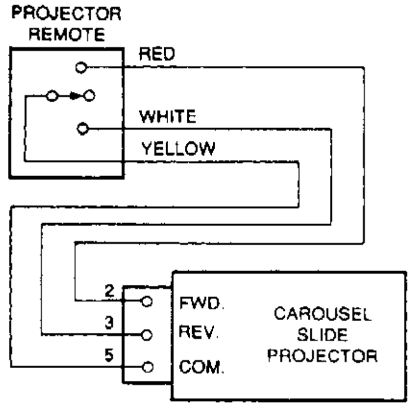


Fig. 1—ORIGINAL CONFIGURATION of the slide projector's remote control.

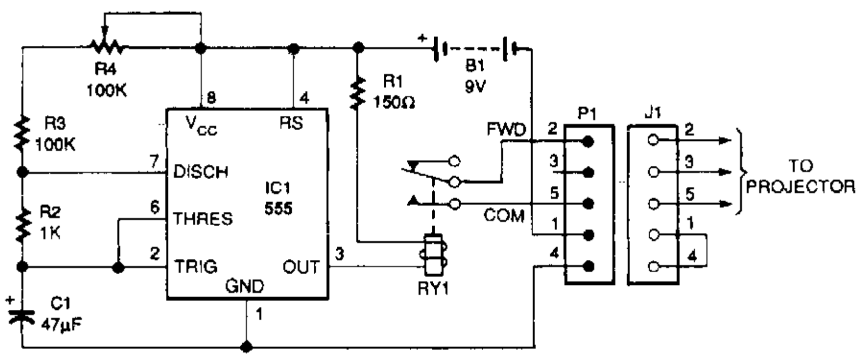


Fig. 2—SCHEMATIC DIAGRAM. The stepper circuit replaces the remote and will automatically advance the slides with a variable time delay.

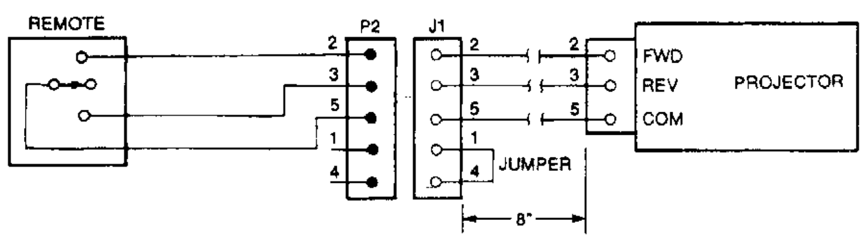
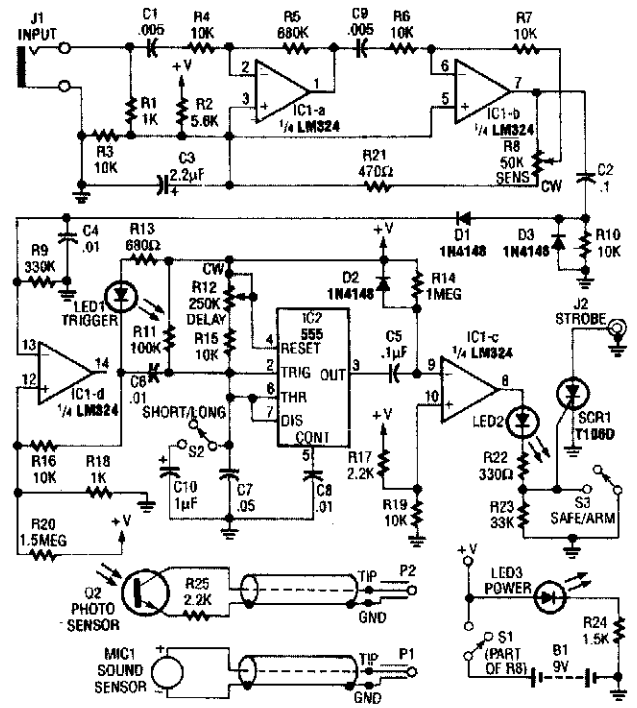


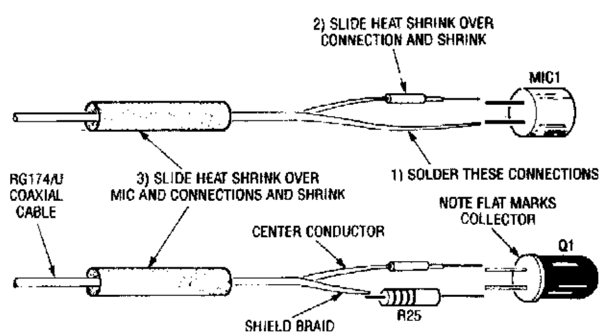
Fig. 3—CABLE MODIFICATION. This will allow the remote and the stepper circuit to be swapped easily.

ELECTRONICS NOW ***Fig. 65-2***

This stepper circuit replaces remote controls and will automatically advance slides in a projector. The time delay is variable with R4. The cable connections are for a Kodak carousel slide projector.

PHOTO SUPER STROBE





MAKE THE SENSOR ASSEMBLIES with heat-shrink tubing and small diameter coaxial cable such as RG-174/U. The space between the coaxial cable and the outer heat-shrink tubing is filled with a little silicone rubber.

ELECTRONICS NOW! ***Fig. 65-3***

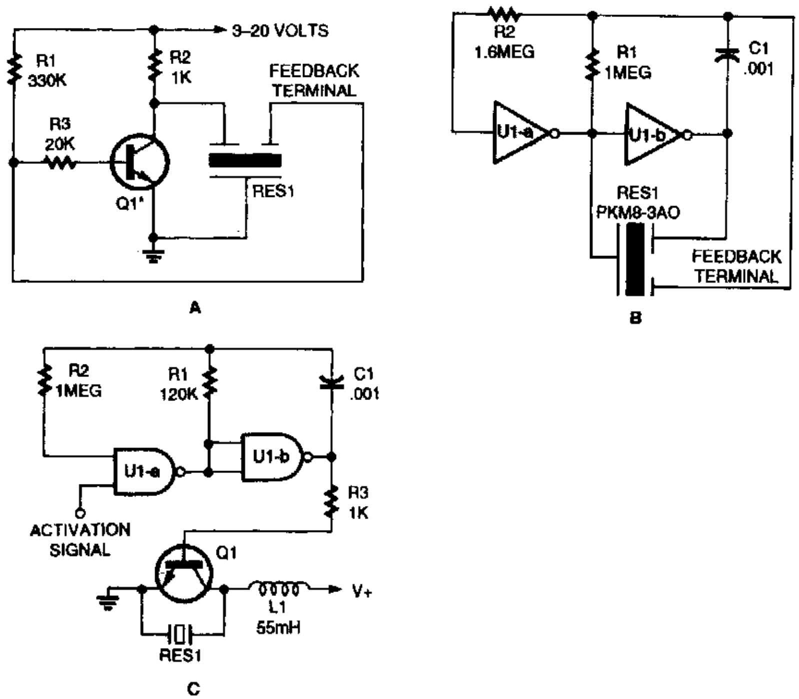
A change in audio or light level on the sensor connected to J1 is amplified by IC1-a and IC1-b (rectified), and used to trigger IC2. R12 sets the delay between the trigger and the flash. IC1-c drives indicator LED2 and triggers SCR1, which sets off the strobe connected to J2. A photo cell or a microphone can be used as a sensor.

66

Piezo Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

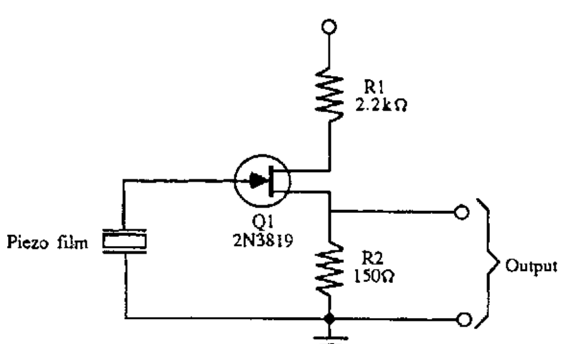
PIEZOELECTRIC DRIVER CIRCUIT



POPULAR ELECTRONICS **Fig. 66-1**

Three-terminal piezoelectric elements are typically driven by transistor circuits (A), or logic gates (B). Two-terminal devices can be driven by two NAND gates. A booster coil is used to compensate for the sound-pressure attenuation caused by the case.

PIEZOELECTRIC BUFFER



R1	2.2 kΩ
R2	150 Ω
Q1	2N3819 FET transistor
Misc.	Piezo disc

All resistors are 5 to 10 percent tolerance, 1/4 watt.

McGRAW-HILL **Fig. 66-2**

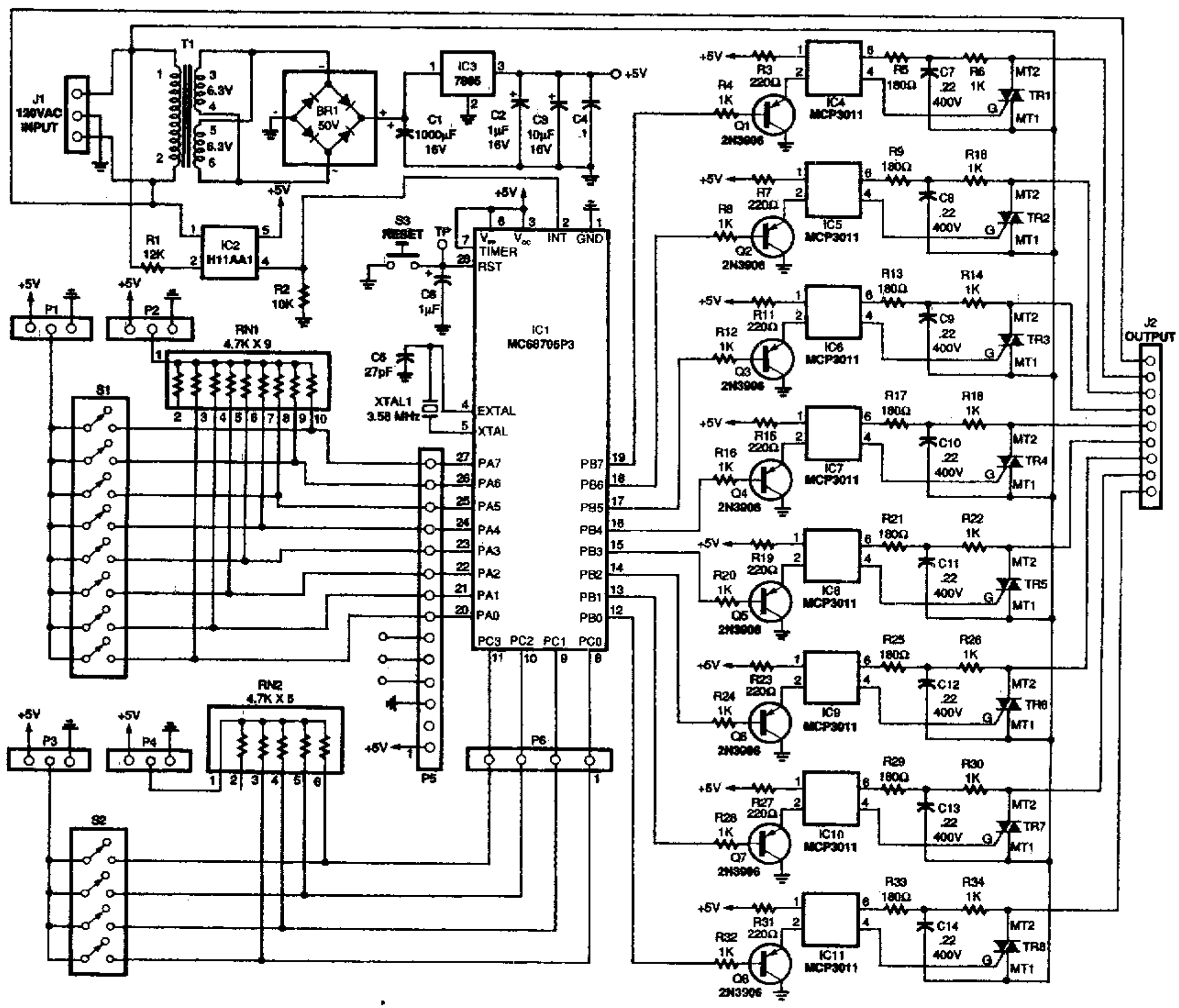
This circuit will serve as a buffer for experiments with Kynar film, a piezoelectric material, or with piezo devices.

67

Power Line Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ac POWER CONTROLLER



ELECTRONICS NOW

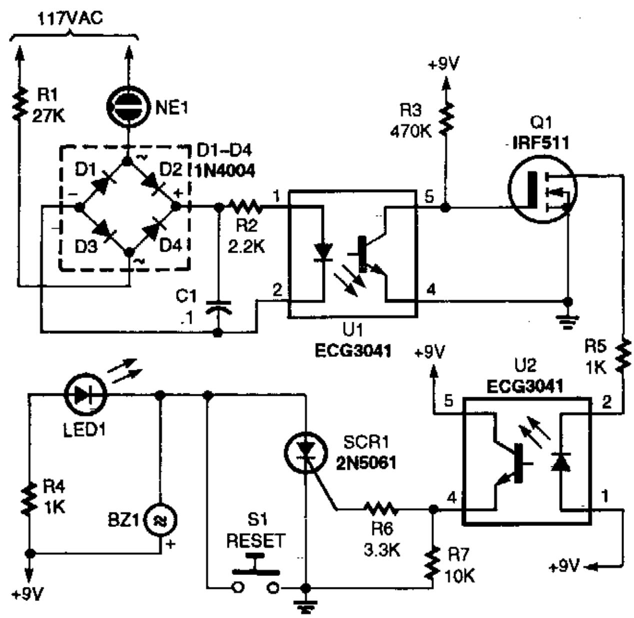
Fig. 67-1

This circuit is used to vary the power delivered to a 120-Vac load under software control. A 68705 micro controller can control eight discrete power triacs, each of which delivers power in 32 smoothly graduated steps, ranging from 0 to 97% of full power. The value delivered to one channel is independent of the value delivered to any other channel. Loads can include light displays, universal motors, heaters, and other appliances.

The power level is set by software, not a potentiometer. The software includes a basic set of routines for processing interrupts and setting the power level. The software also includes five test and demonstration routines for putting the circuit through its paces. Moreover, there's plenty of room to add your own routines to the 68705's built-in EPROM.

The basic circuit is simple, yet versatile enough to accept inputs from on-board DIP switches; alternatively, the inputs can be driven from a microcomputer bus or parallel port, or a stand-alone device with TTL-compatible outputs. There are 12 input bits to set modes and specify values.

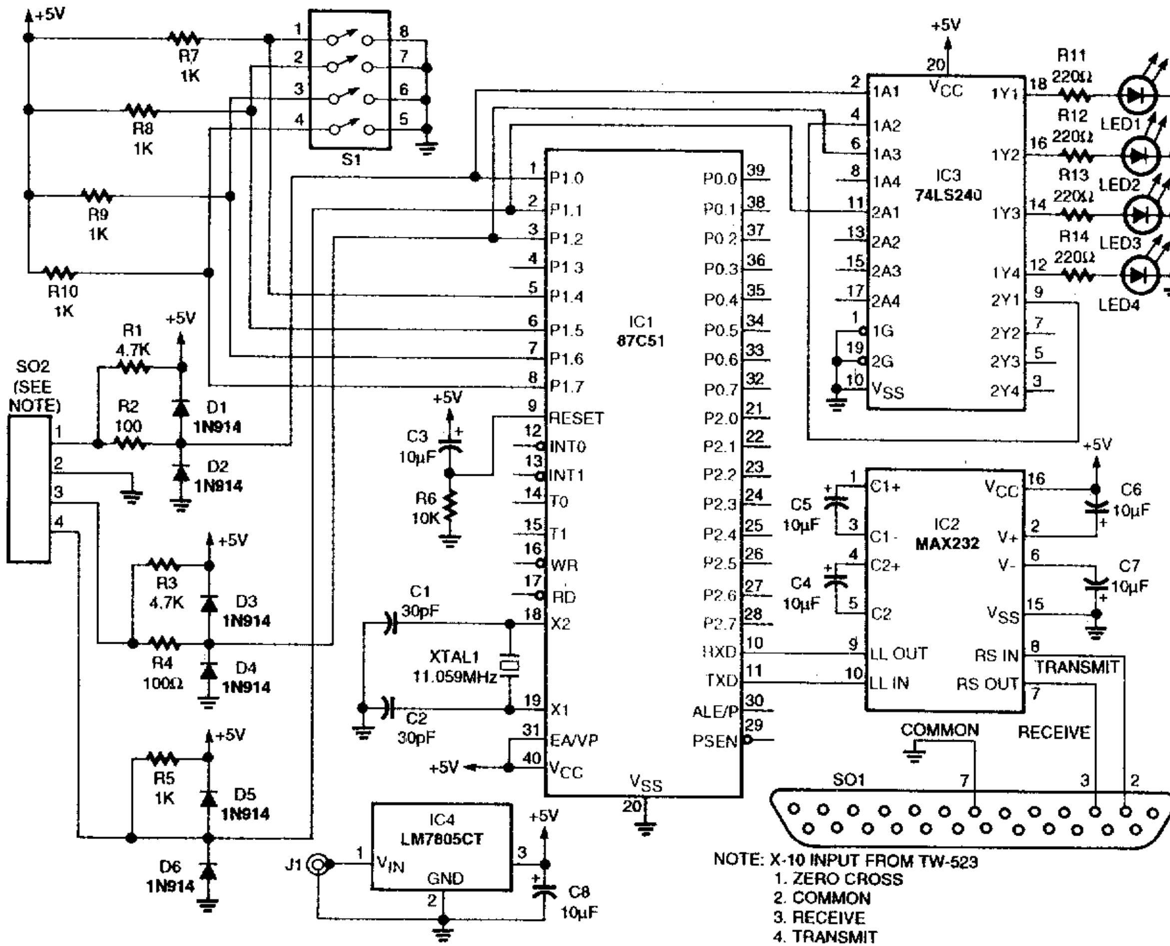
ac POWER-LINE MONITOR



POPULAR ELECTRONICS *Fig. 67-2*

When the power-line voltage source fails, Q1 turns on, activates optoisolator U2, and triggers SCR1. For small SCRs, U1 might directly trigger SCR1.

POWER-LINE MODEM FOR COMPUTER CONTROL

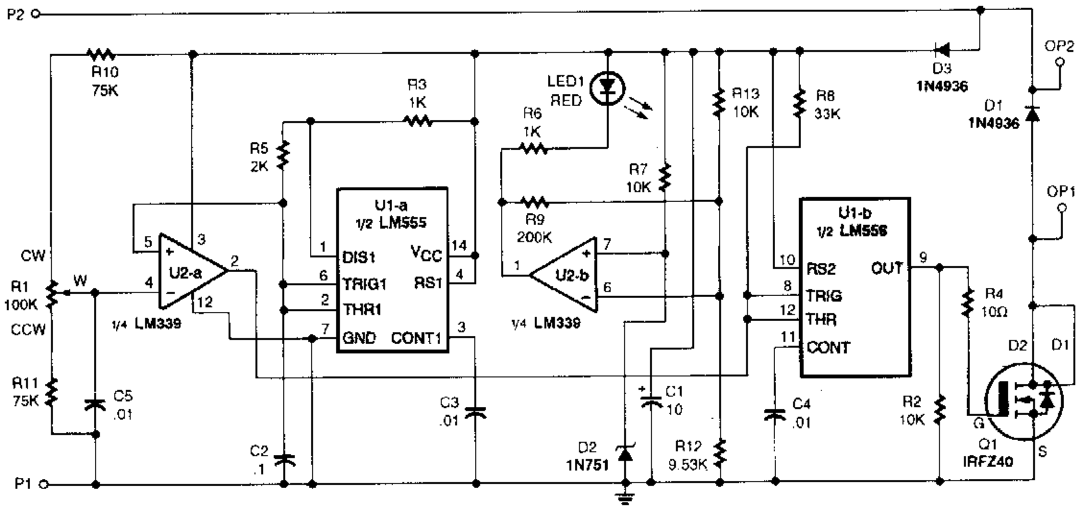


ELECTRONICS NOW

Fig. 67-3

This circuit uses an 87C57 microcontroller and a few peripherals to condition X-10 power-line carrier-code formats from a personal computer to use an X-10 power-line interface in a home-control system. Software details are available in the reference.

LOW-VOLTAGE POWER CONTROLLER



POPULAR ELECTRONICS *Fig. 67-4*

The circuit has a duty-cycle generator that will produce an output varying from fully off to fully on and pulses of any duty cycle in between the two extremes.

This method of operation is called *PWM (pulse width modulation)*. The circuit can be fed from any dc supply source of between 10 to 15 V. Half of an LM556 dual oscillator/timer and U2-a (¼ of an LM339 quad comparator) combine to form a voltage-to-pulse-width converter. The first half of the dual oscillator/timer (U1-a) is configured as an astable oscillator, generating a continuously oscillating ramp voltage. Op amp U2-a compares the voltage at its noninverting input (pin 5)—which is connected to pins 2 and 6 of U1-a—to the voltage at its inverting input (pin 4). The op amp will produce a low output if R1's wiper voltage is higher than the instantaneous voltage that is present at pins 2 and 6 of U1-a. The output of U2-a at pin 2 will have an on/off ratio that is proportional to the voltage at R1's wiper.

The output of U2-a is fed to U1-b, which is used to buffer the signal. The low-impedance, pulsed output of U1-b at pin 9 is fed to the gate of MOSFET Q1, driving it on or off. The circuit also has a power-input detector, built around U2-band and LED1. If the input power is OK, LED1 will shut off.

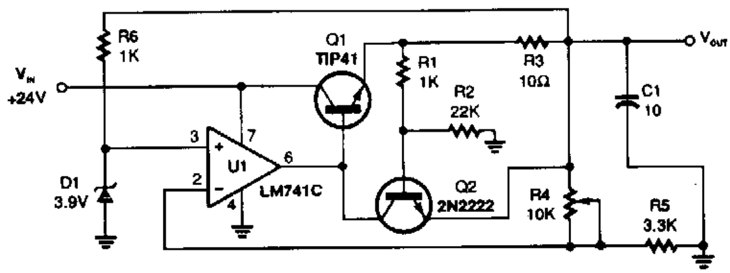
Diode D1 is used to suppress the reverse voltage spikes that are generated by inductive loads during turn off; without that diode, the MOSFET might be destroyed. If the circuit will not be used to drive inductive loads (motors), D1 can be eliminated.

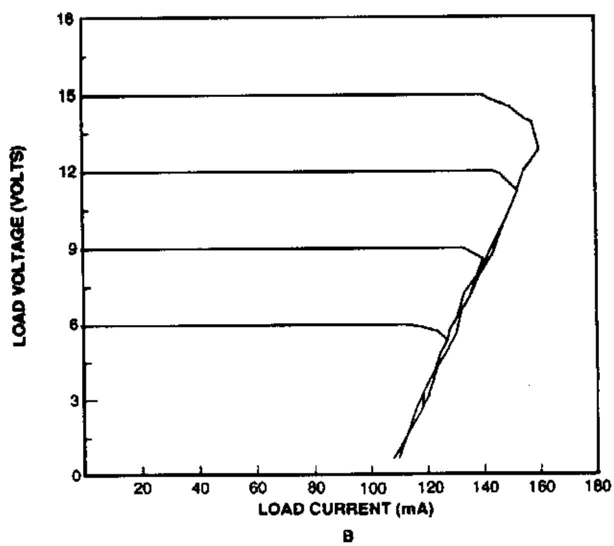
68

Power Supply Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

FOLDBACK CURRENT LIMITER

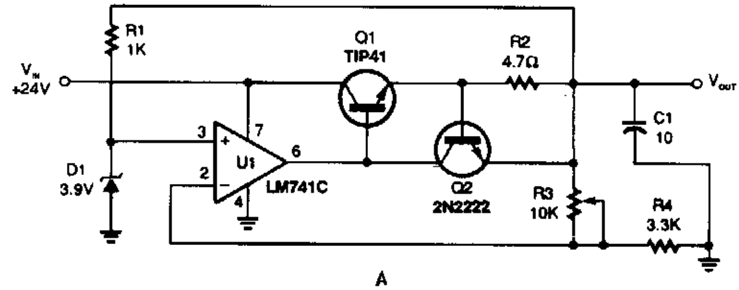




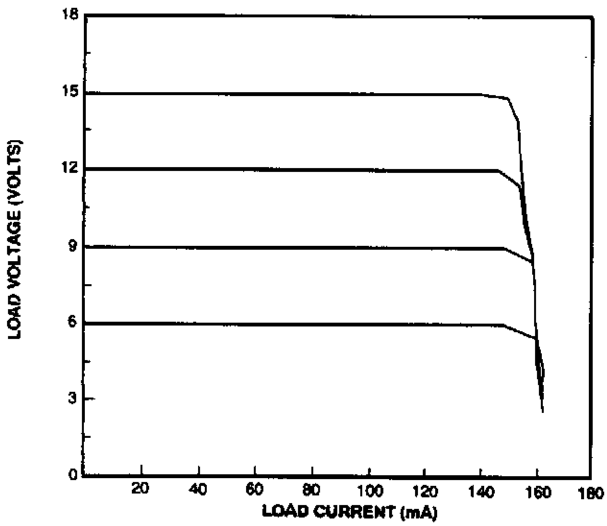
POPULAR ELECTRONICS *Fig. 68-1*

This regulator uses the drop across R3 to sense current draw, turning on Q2, removing drive from Q1, and lowering the output voltage. Limiting occurs when Q2 has 0.65 V across the base-emitter junction. This circuit has foldback characteristics as seen from the figure.

CURRENT-LIMITING REGULATOR CIRCUIT



A



B

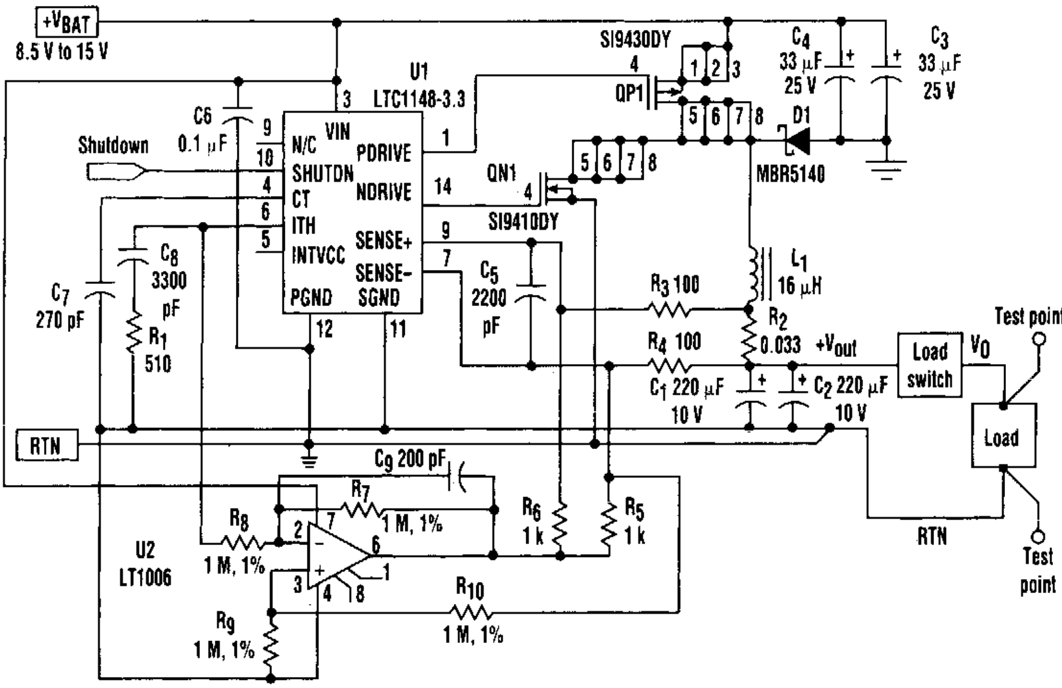
As shown in B, maximum load current is practically the same at all supply voltages with constant current limiting.

POPULAR ELECTRONICS ***Fig. 68-2***

This regulator uses the drop across R2 to turn on Q2, removing base drive from Q1 and reducing the current through R1. About 0.65 V must be dropped across R2 before limiting occurs. To set limit current,

$$R_2 \approx \frac{0.65}{I_{\text{LIMIT (amps)}}}$$

$$\textit{Output voltage} = V_{\text{OUT}} = (3.9)\,\frac{(R_3 + R_4)}{R_4}$$



ELECTRONIC DESIGN

Fig. 68-3

In many switching-regulator applications for portable computers, the microprocessor is located some distance from the power supply. With the latest processors, total load currents range into several amperes. Thus, regulation at the load can become a problem.

The I_{TH} pin (pin 6) of the LTC1148 is approximately proportional to the load current. It scales nearly linearly from 0 V at no load to 2.0 V at current limit. U2, acting as a unity-gain differential amplifier, inverts the U1 pin 6 voltage (referenced to SENSE—, pin 7) and causes a current proportional to load current to flow in resistors R5 and R6. A small voltage drop appears across current-sense filter resistors R3 and R4. This makes the voltage measured by the internal feedback divider appear low. The duty factor is adjusted to bring this back to the correct voltage. As a result, the output is increased slightly as a function of load. Capacitor C9 rolls off the high-frequency gain of the correction amplifier.

$$R_{comp} = \frac{(V_{pin6} \times R_{filt})}{V_{corr}}$$

where: $R_{comp} = R_5 = R_6$
$R_{filt} = R_3$ and R_4
V_{corr} = *Measured drop.*

TRANSFORMERLESS dc POWER SUPPLY I

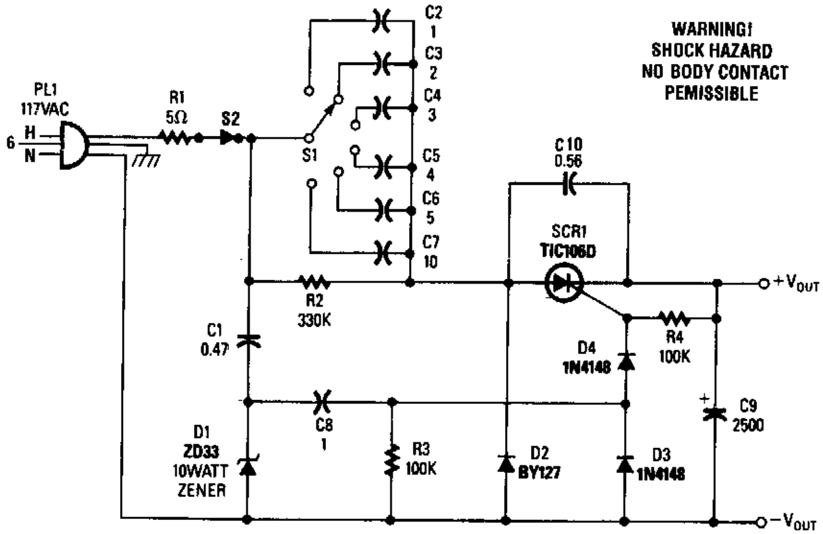


TABLE 1—OUTPUT CURRENT/VOLTAGE AT SPECIFIED LOADS

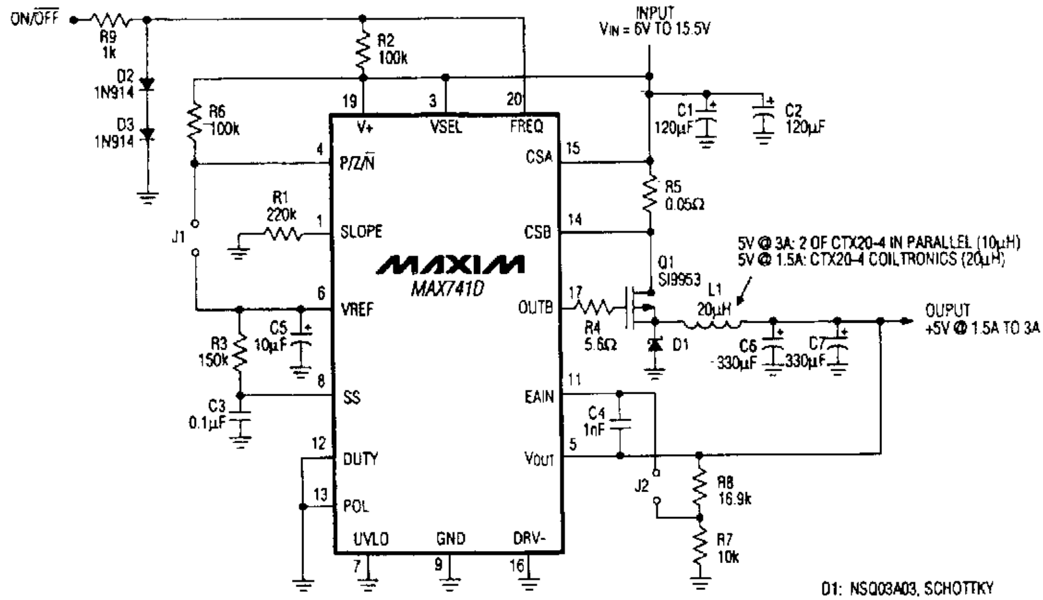
Capacitance (μF)	Load					
	100 Ohms		200 Ohms		1000 Ohms	
	V_{OUT} (V)	Current (mA)	V_{OUT} (V)	Current (mA)	V_{OUT} (V)	Current (mA)
1	3.2	31	6.0	29	25	24
2	6.4	61	11.2	54	41	41
3	9.0	87	16.1	78	52	52
4	11.8	113	20.7	100	61	61
5	15.5	147	24.7	120	67	67
6	17.8	169	28.8	140	68	68
7	18.5	176	31.9	155	69.4	68
8	20.3	195	36.8	173	70	71
9	22.8	220	41.0	193	70	71
10	24.9	238	42.0	204	71	71
11	27.1	259	44.9	219	—	—
14	33.0	317	52.7	257	—	—
20	43.5	422	65.8	322	—	—

POPULAR ELECTRONICS

Fig. 68-4

An SCR fires on the positive half cycles of the ac line voltage. Switched capacitors are used to select the output voltage. These must all be ac-rated, nonpolarized types.

+5-V AT 1.5- TO 3-A SUPPLY, +6- TO +15-V INPUT

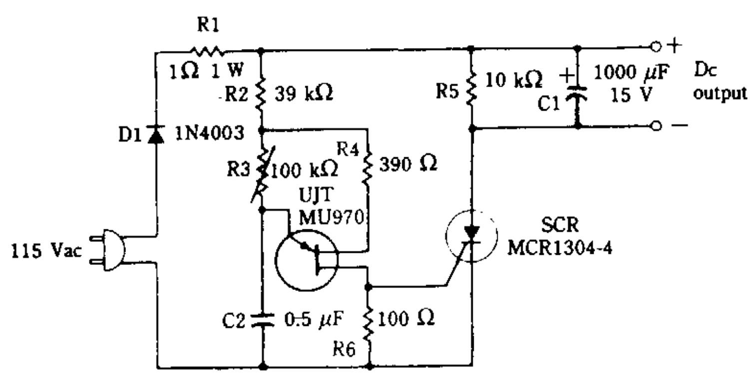


MAXIM

Fig. 68-5

Operating efficiencies of 80 to 90% are possible using the MAX741D and this circuit.

TRANSFORMERLESS dc POWER SUPPLY II

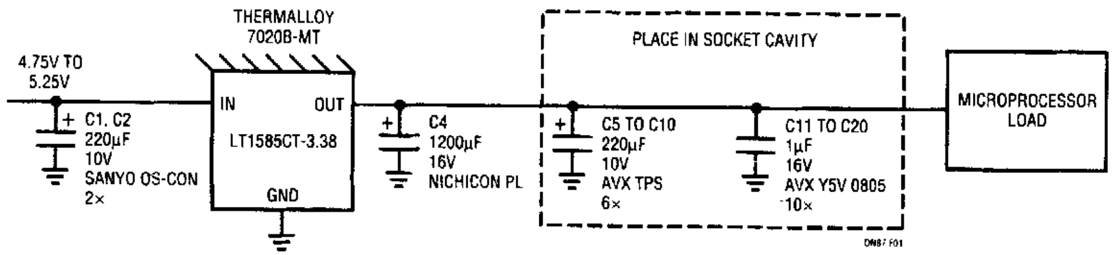


McGRAW-HILL

Fig. 68-6

Although it is simple, this supply can provide 10 to 15 V at 100 mA directly from the ac lines. This circuit has no isolation from the ac line; therefore, there is a shock hazard and it should only be used where no possibility of contacting external devices, circuits, or personnel exists.

FAST 3.3-V REGULATOR



AVX CORPORATION: (803) 448-9411
NICHICON (AMERICA) CORPORATION: (708) 843-7500
SANYO VIDEO COMPONENTS (USA) CORPORATION: (619) 661-6322
THERMALLOY INCORPORATED: (214) 243-4321
FOR CORRECT OPERATION OF MICROPROCESSOR, DO NOT SUBSTITUTE COMPONENTS

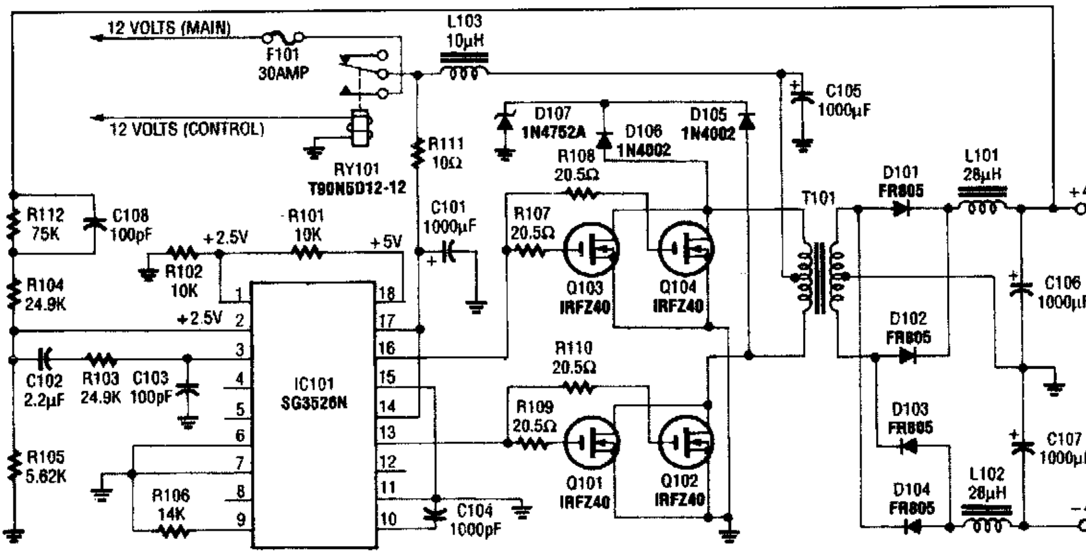
LINEAR TECHNOLOGY ***Fig. 68-7***

New high-performance microprocessors require a fresh look at power-supply transient response. The LT1585 linear regulator features 1% initial accuracy, excellent temperature drift and load regulation, and virtually perfect line regulation. Complementing superb dc characteristics, the LT1585 exhibits extremely fast response to transients. Transient response is affected by more than the regulator itself. Stray inductances in the layout and bypass capacitors, as well as capacitor ESR dominate the response during the first 400 ns of transient.

The figure shows a bypassing scheme developed to meet all the requirements for the Intel P44C-VR microprocessor. Input capacitors C1 and C2 function primarily to decouple load transients from the 5-V logic supply. The values used here are optimized for a typical 5-V desktop computer "silver box" power-supply input. C5 to C10 provide bulk capacitance at low ESR and ESL, and C11 to C20 keep the capacitance at low ESR and ESL low at high (>100 kHz) frequencies. C4 is a damper and it minimizes ringing during setting. Trace C is the load current step, which is essentially flat at 4 A with a 20-ns rise time.

Trace A is the output settling response at 20 mV per division. Cursor trace B marks –46 mV relative to the initial output voltage. At the onset of load current, the microprocessor socket voltage dips to –38 mV as a result of inductive effects in the board and capacitors, and the ESR of the capacitors. The inductive effects persist for approximately 400 ns. For the next 3 µs, the output droops as the load current drains the bypass capacitors. The trend then reverses as the LT1585 catches up with the load demand, and the output settles after approximately 50 µs. Running 4 A with a 1.7-V drop, the regulator dissipates 6.8 W.

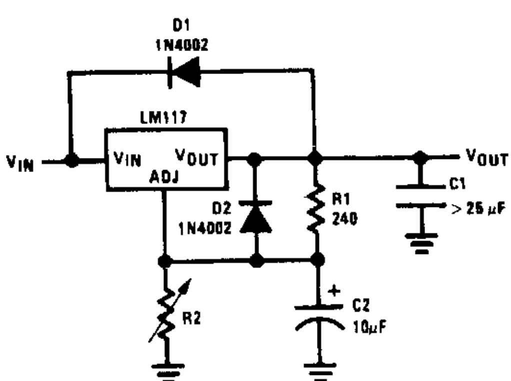
POWER SUPPLY FOR HIGH-POWER AUTOSOUND AMP



RADIO-ELECTRONICS *Fig. 68-8*

A switching mode inverter is used with a pulse-width modulation voltage regulator (SG35260). Four IRF240 power MOSFETs are used as switches. The output is ±47 V at about 5 A peak. Transformer T101 is a four-turn center tapped primary, and 16-turn center tapped secondary on a Ferrox cube ETD-34 core.

IC REGULATOR PROTECTION



$V_{OUT} = 1.25V\left(1 + \frac{R2}{R1}\right) + R2 \bullet I_{ADJ}$

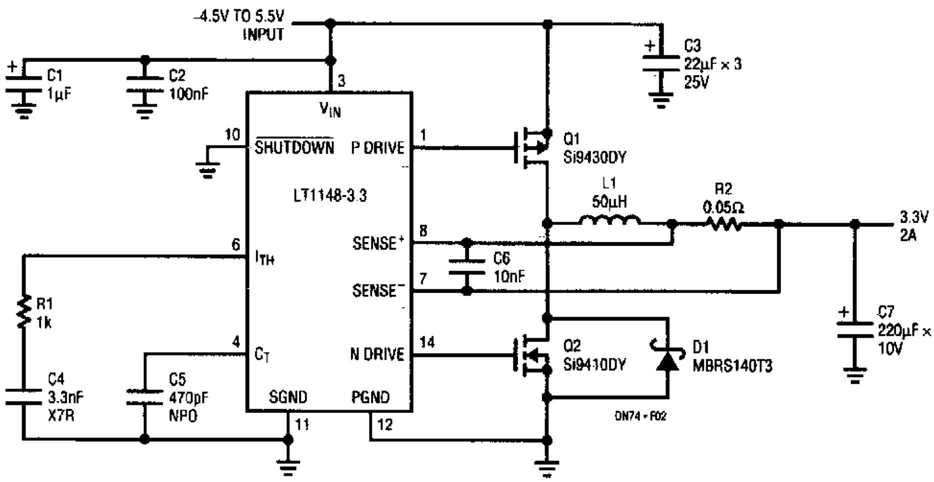
D1 protects against C1 (input shorts)

D2 protects against C2 (output shorts)

NATIONAL SEMICONDUCTOR *Fig. 68-9*

This circuit protects an IC regulator against various fault conditions.

3.3-V SWITCHING REGULATOR

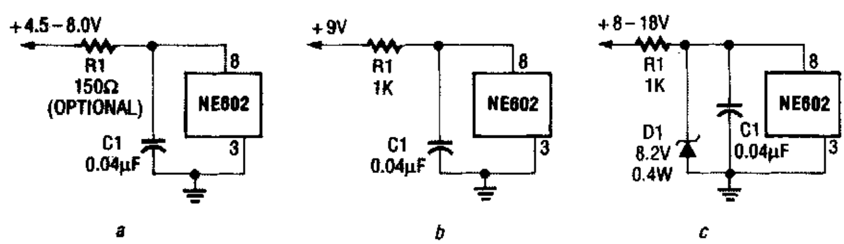


LINEAR TECHNOLOGY

Fig. 68-10

For the LT1129-3.3, dissipation amounts to a little under 1.5 W at full output current. The 5-lead surface-mount DD package handles this without the aid of a heatsink, provided that the device is mounted over at least 2500 mm^2 of ground or power-supply plane. Efficiency is around 62%; dissipation in linear regulators becomes prohibitive at higher current levels, where they are supplanted by high-efficiency switching regulators. The synchronous buck converter is implemented with an LTC1148-3.3 converter. The LTC1148 uses both Burst Mode™ operation and continuous, constant off-time control to regulate the output voltage, and maintain high efficiency across a wide range of output loading conditions.

NE602 POWER-SUPPLY OPTIONS

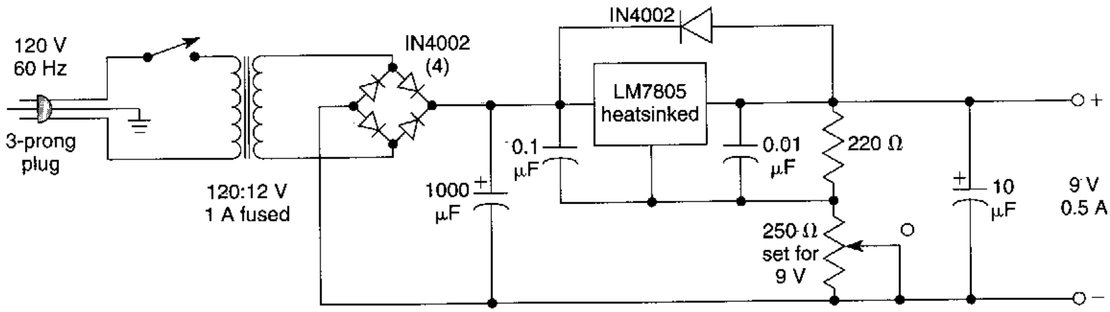


RADIO-ELECTRONICS

Fig. 68-11

Here, Figs. 68-11A through 68-11C show an RC-filter used as both current limiter (R1) and integrator (C1), as well as for isolation. In Fig. 68-11A, +4.5 to 8.0 Vdc is the normal operating range of the NE602. In Fig. 68-11B, R1 drops voltage, and is used because a +9-V battery can go higher, and a +9-V wall supply can produce up to 11 V. In Fig. 68-11C, a +8- to 18-Vdc supply is regulated using a 8.2-V Zener for D1.

SIMPLE 9-V SUPPLY

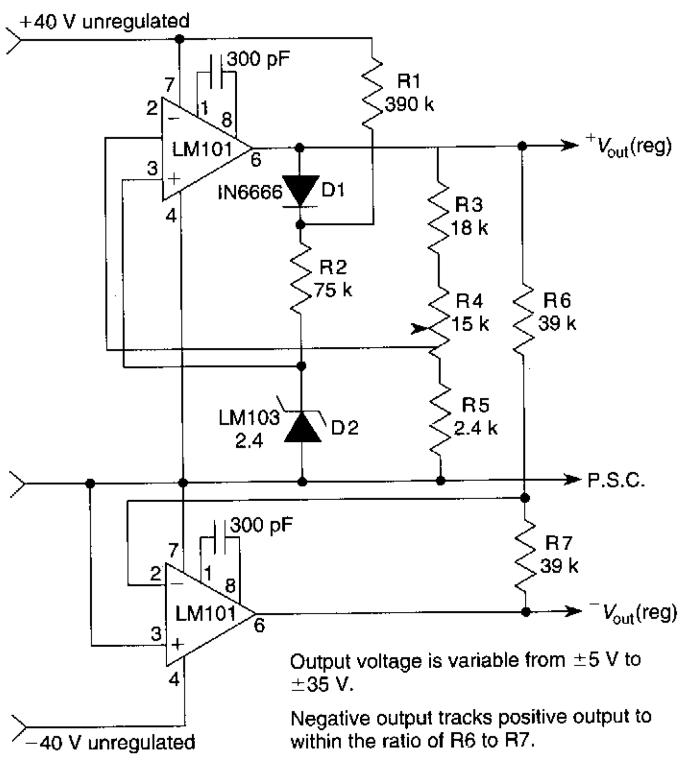


WILLIAM SHEETS

Fig. 68-12

This circuit uses an LM7805 with a resistive voltage divider in the common leg of the regulator. The regulator can be "fooled" into producing an apparent higher output voltage in this manner. This supply is useful for running radios, tape recorders, or other 9-V devices.

TRACKING POWER SUPPLY

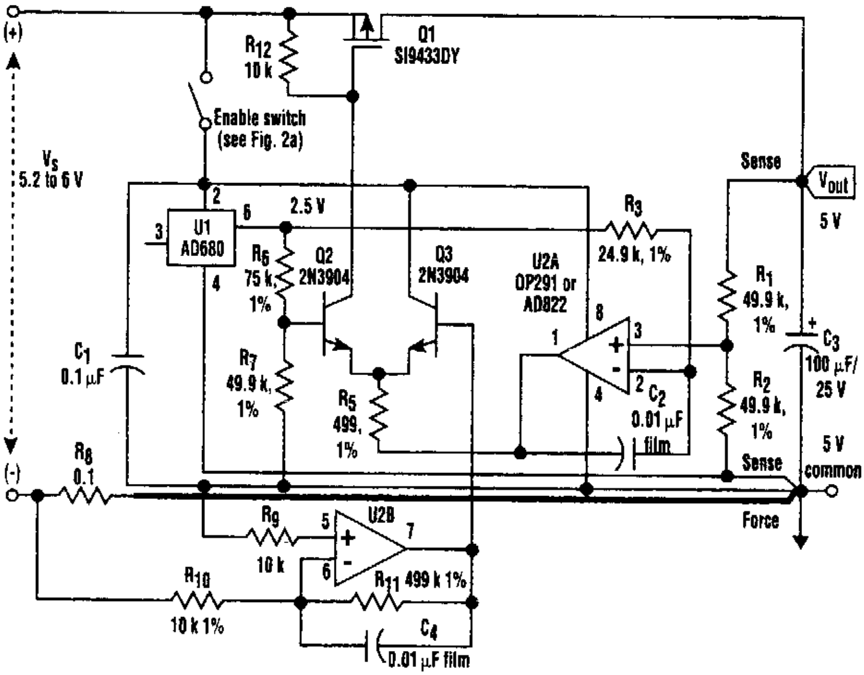


NATIONAL SEMICONDUCTOR

Fig. 68-13

Two op amps are used in this basic op-amp regulator circuit. The outputs can be fed to current amplifier stages or emitter followers, if needed.

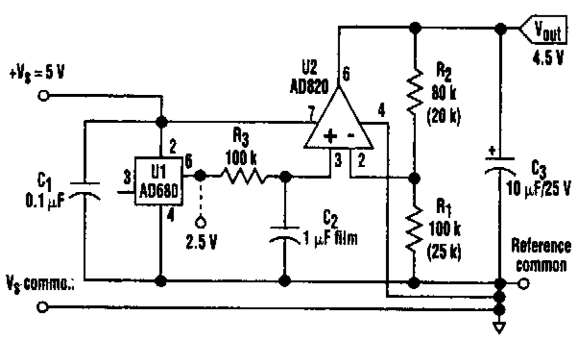
POWER EFFICIENT VOLTAGE REGULATOR



ELECTRONIC DESIGN *Fig. 68-14*

Included in the many features of this power-efficient, voltage-regulator circuit is shutdown power control with a current output up to several hundred milliamperes (expandable to amperes, if desired). Current limiting can be preset to a fixed level for controlled dissipation in Q1 and the circuit requires no auxiliary voltage supply for the pass transistor.

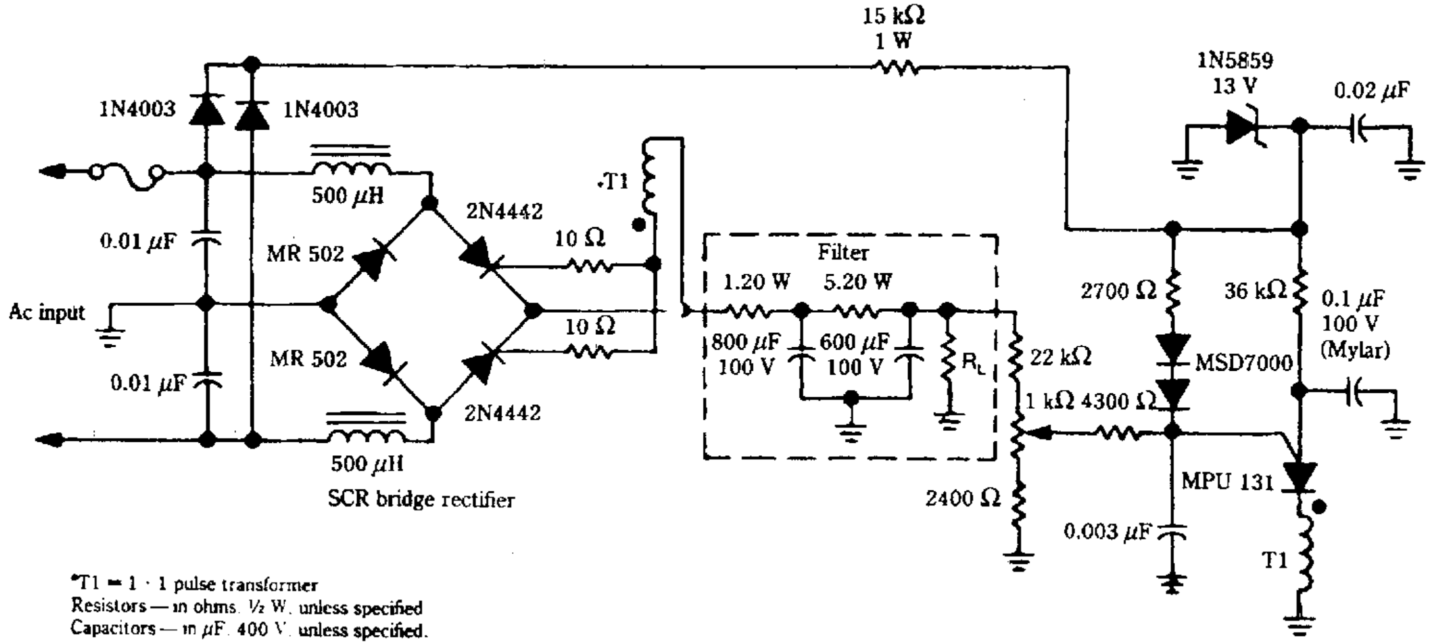
LOW DROP-OUT REGULATOR



ELECTRONIC DESIGN *Fig. 68-15*

This low-dropout reference produces a 4.5-V output from a supply just a few hundred millivolts greater. With 1-mA dc loading, it maintains a stable 4.5-V output for inputs down to 4.7 V.

SCR SWITCHING SUPPLY FOR COLOR TV RECEIVERS

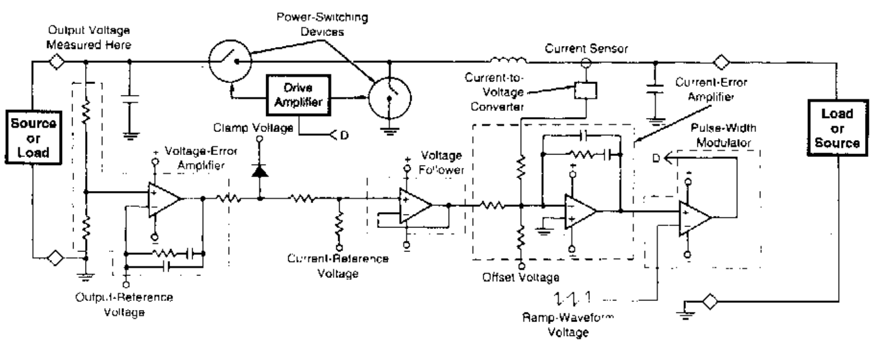


McGRAW-HILL

Fig. 68-16

An output +80 V at up to 1.5 A is available from this supply. A minimum load of 200 mA is required because of the SCR holding current. Notice that no ac line isolation is provided and a shock hazard exists.

REGULATOR CIRCUIT FOR BILATERAL SOURCE/LOAD POWER SYSTEM



NASA TECH BRIEFS

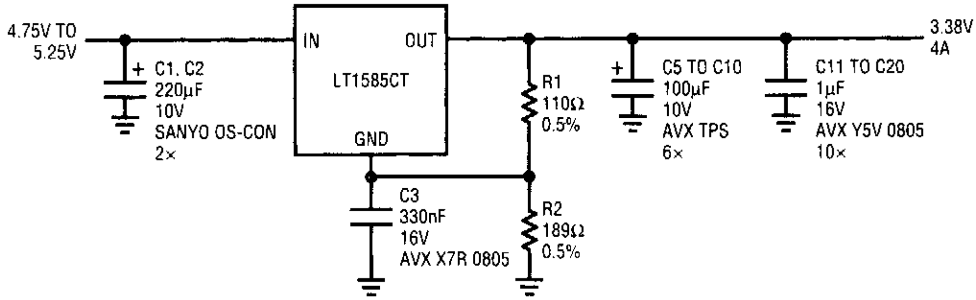
Fig. 68-17

The figure shows a circuit that regulates an output voltage, regardless of the direction of flow of output current. More specifically, it regulates the voltage at the left source or load, which can either supply power to or receive power from the right load or source, depending on the voltages and the direction of flow of current at the load/source terminals.

The overall system can be characterized as a voltage-controlled current source with bilateral current capability. The current flowing between the two source/loads, averaged over a power-switching cycle, is made to depend on the pulse-width modulation that governs the operation of the two power-switching devices, and this pulse-width modulation is, in turn, a function of amplified current-error and voltage-error signals. The voltage error is the difference between the actual output voltage and the output-reference voltage, which is the nominal output voltage at zero current. The pulse-width modulation is varied to increase or decrease the current, as needed, to limit the excursion of output voltage from the reference value.

An additional feature of this control circuit is that the maximum current in either direction can be limited by limiting the excursion of the output voltage from the zero-current value. Thus, external current-limiting circuitry is not necessary.

FAST 3.3-V ADJUSTABLE REGULATOR



LINEAR TECHNOLOGY ***Fig. 68-18***

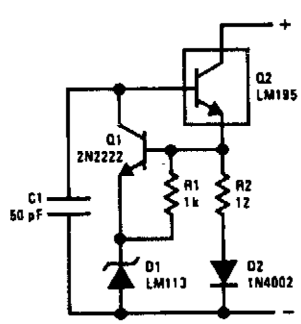
The adjustable version of the LT1585 makes it relatively easy to accommodate multiple microprocessor power-supply voltage specifications. To retain the tight tolerance of the LT1585 internal reference, a 0.5% resistor adjustment is recommended. R1 is sized to carry approximately 10 mA idling current (≤124 Ω), and R2 is calculated from:

$$R_2 = \frac{V_o - V_{ref}}{\frac{V_{ref}}{R_1}} + 1_{ADJ}$$

where:

$$1_{ADM} = 60\ \mu \text{ and } V_{ref} = 1.250\text{ V.}$$

TWO-TERMINAL 100-mA CURRENT REGULATOR

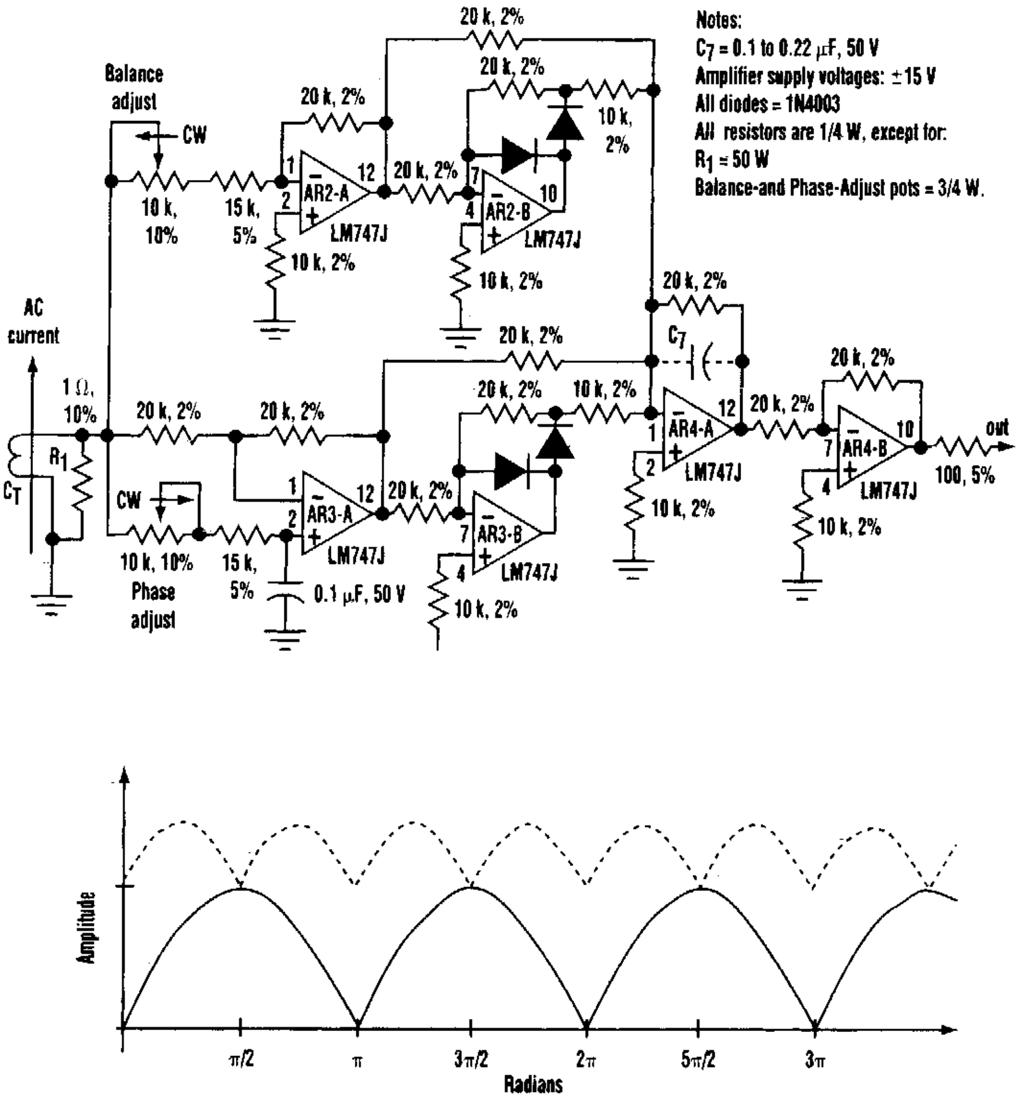


NATIONAL SEMICONDUCTOR ***Fig. 68-19***

The circuit has a low temperature coefficient and operates down to 3 V. The reverse base current of the LM195 biases the circuit.

A 2N2222 is used to control the voltage across current-sensing resistor, R2 and diode D1, and therefore the current through it. The voltage across the sense network is the V_{BE} of the 2N2222 plus 1.2 V from the LM113. In the sense network, R2 sets the current and D1 compensates for the V_{BE} of the transistor. Resistor R1 sets the current through the LM113 to 0.6 mA.

TWO-PHASE RECTIFIER

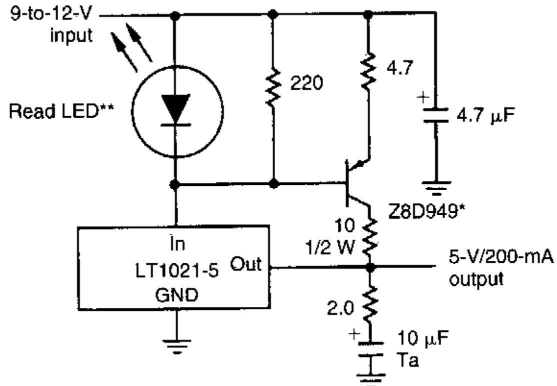


A single-phase AC signal can be converted to two phase with this circuit. It rectifies and sums the signal to a dc voltage level.

ELECTRONIC DESIGN *Fig. 68-20*

The waveform generated by the two-phase rectifier illustrates that the ripple is less than half that of a conventional single-phase circuit's waveform. Also, the ripple frequency is double that of the conventional circuit. The circuit will follow amplitude changes in the ac input signal very rapidly, and it works equally well with current or voltage inputs.

LOW-NOISE 5-V SUPPLY



*Zetex Inc.
**Glows in current limit. Do not omit.

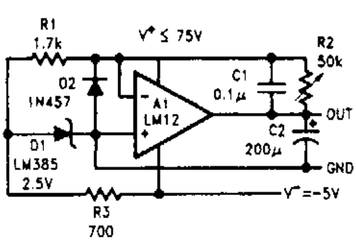
ELECTRONIC DESIGN *Fig. 68-21*

Standard three-terminal regulator ICs can be noisy. The key is the noise over the 10-Hz to 10-kHz band; measurements revealed a 40-dB improvement over standard three-terminal regulators.

The regulator is built around a 5-V buried-Zener reference. It's the buried Zener's inherently low noise that makes the finished supply so quiet. Measured over a 10-Hz to 10-kHz band, the 5-V output contains just 7 μV rms of noise at full load. The 10-Hz to 10-kHz noise can be further reduced to 2.5 μV rms by adding a 100-μH, 1000-μF output filter. The noise characteristics of the reference are tested and guaranteed to a maximum of 11 μV over the band of interest.

An external boost transistor, the ZBD949, provides gain to meet a 200-mA output current requirement. Current limiting is achieved by ballasting the pass transistor and clamping the base drive. Although the oscillator only requires 200 mA, it's possible to extend the output current to at least 1 A.

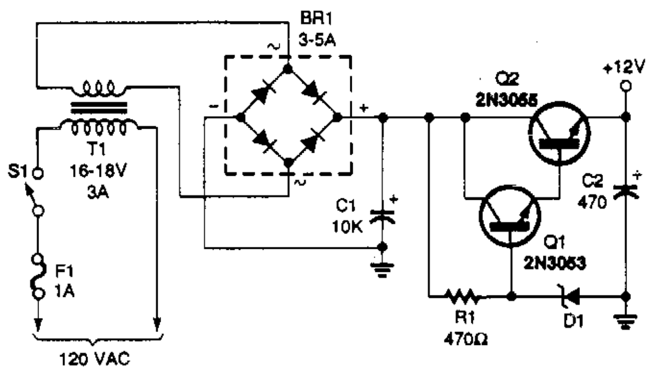
POSITIVE REGULATOR WITH 0- TO 70-V OUTPUT



NATIONAL SEMICONDUCTOR *Fig. 68-22*

The op amp has one input at ground and a reference current drawn from its summing junction. With this arrangement, the output voltage is proportional to setting resistor R2. A negative supply is used to operate the op amp within its common-mode range, providing zero output with sink current and power a low-voltage bandgap reference, D1. The current drawn from this supply is under 150 mA, except when sinking a load current. The output load capacitor, C2, is part of the op-amp frequency compensation.

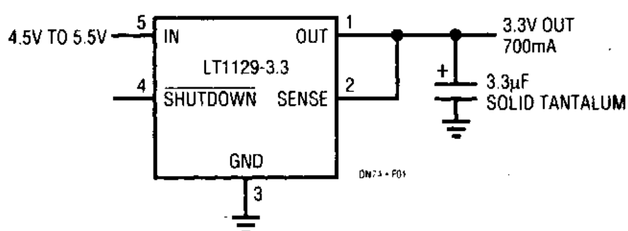
SIMPLE 12-V POWER SUPPLY



POPULAR ELECTRONICS *Fig. 68-23*

This 12-V power supply is easy-to-build, and it produces a smooth output. D1 is a 14-V, ½-W Zener diode. The voltage can be varied by a few volts up or down to change the output voltage.

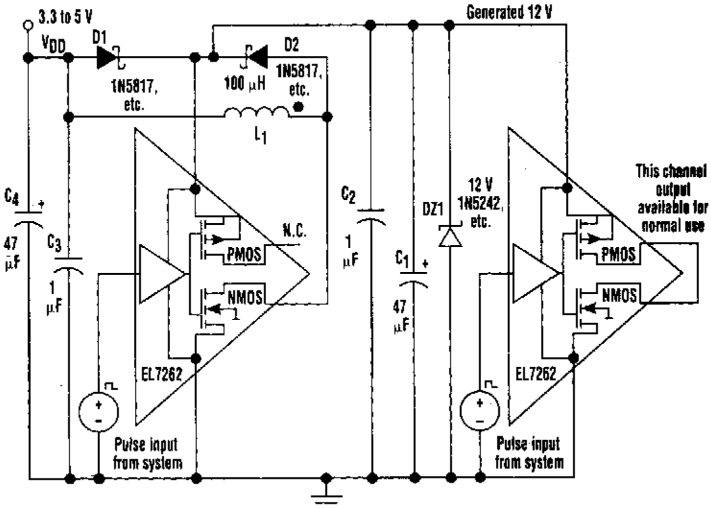
3.3 V FROM 5-V LOGIC SUPPLIES



LINEAR TECHNOLOGY *Fig. 68-24*

Microprocessor chip sets and logic families that operate from 3.3-V supplies are gaining acceptance in both desktop and portable computers. Computing rates, and in most cases, the energy consumed by these circuits, show a strong improvement over 5-V technology. The main power supply in most systems is still 5 V, necessitating a local 5-V to 3.3-V regulator. Linear regulators are viable solutions at lower ($I_o \leq 1$ A) currents, but they must have a low dropout voltage in order to maintain regulation with a worst-case input of only 4.5 V. The figure shows a circuit that converts a 4.5-V minimum input to 3.3 V with an output tolerance of only 3% (100 mV). The LT1129-3.3 can handle up to 700 mA in surface-mount configurations, including both 16-μA shutdown and 50-μA standby currents for system sleep modes. Unlike other linear regulators, the LT1129-3.3 combines both low-dropout and low-voltage operation. Small input and output capacitors facilitate compact, surface-mount designs.

12-V SUPPLY



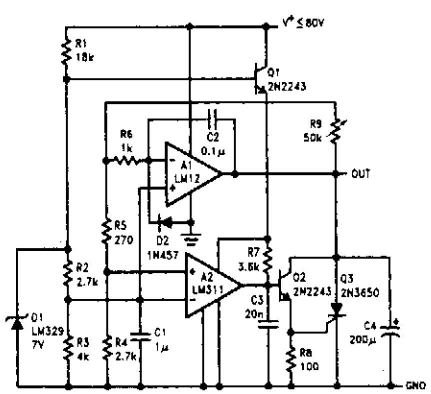
ELECTRONIC DESIGN **Fig. 68-25**

When driving a power MOSFET from a 5-V or 3.3-V system, a significant number of components usually are needed to generate an extra +12 V.

It's possible, however, to apply the second channel in a typical dual MOSFET driver to derive a +12-V power supply. By using a driver with the drains brought to separated pins, you can connect an inductor between the n-channel drain and the logic supply without connecting the p-channel device.

The driver operates as a standard flyback-style switched-mode circuit (see the figure). When the output n-channel device is on, current starts flowing in the inductor, which stores energy. When the n-channel device is turned off, current must continue flowing. Therefore, it flows through diode D2 to charge up C1 and C2. As the cycle repeats, the C1 and C2 voltage rises until the Zener diode prevents further voltage rise. This is needed to prevent the driver's derived supply from exceeding the part's maximum voltage rating.

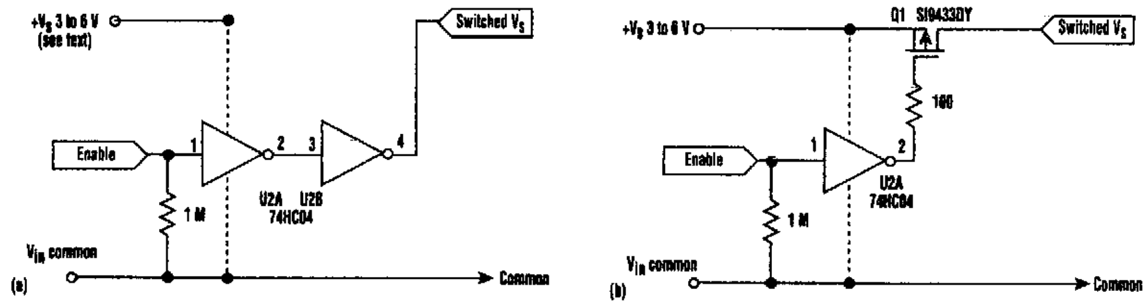
4- TO 70-V REGULATOR



NATIONAL SEMICONDUCTOR **Fig. 68-26**

This regulator operates from a single supply. If the op amp is not able to control an overvoltage condition, the SCR will crowbar the output.

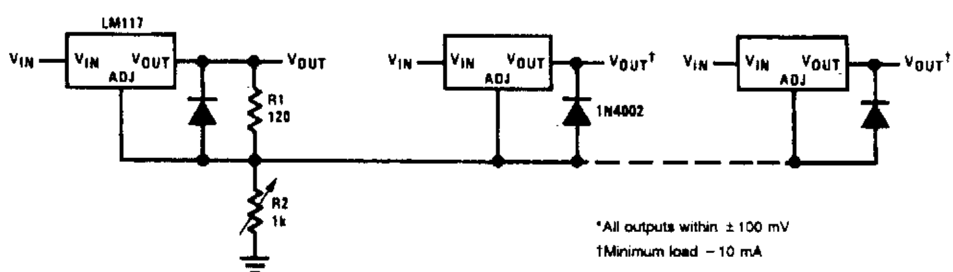
SWITCHED POWER-CONTROL CIRCUITS



ELECTRONICS DESIGN *Fig. 68-27*

Load currents of a few milliamperes to amperes can be turned on and off by these switched power-control circuits. The U2 CMOS inverter stage works as a simple power switch for load currents less than 5 mA (Fig. 68-27A), allowing easy reference shutdown. If appreciably higher switched output currents are called upon, an alternate CMOS inverter driving a low-threshold PMOS device can be used to switch currents of up to 1 A or more (Fig. 68-27B).

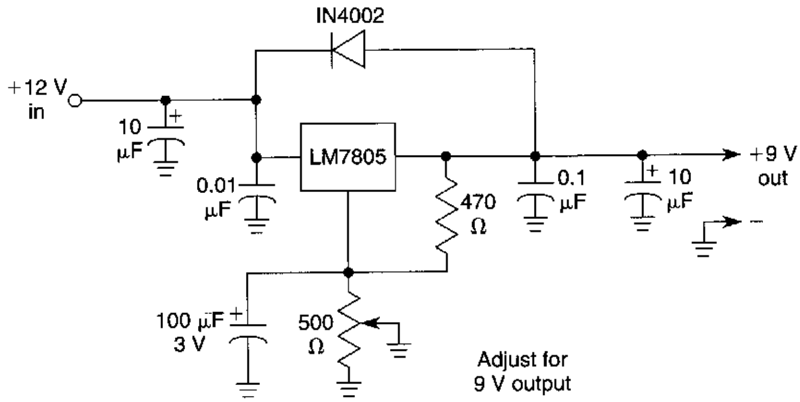
MULTIPLE ON-CARD REGULATOR ADJUSTER



NATIONAL SEMICONDUCTOR *Fig. 68-28*

This circuit allows one pot to control several on-card regulators for adjustment within ±100 mV of each other.

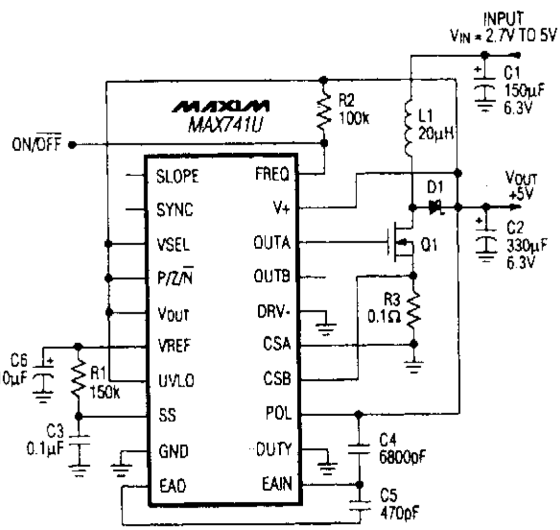
SIMPLE 9-V POWER SUPPLY



WILLIAM SHEETS *Fig. 68-29*

This supply will provide 9-V transistor radios or cassettes from a 12-V auto electrical system.

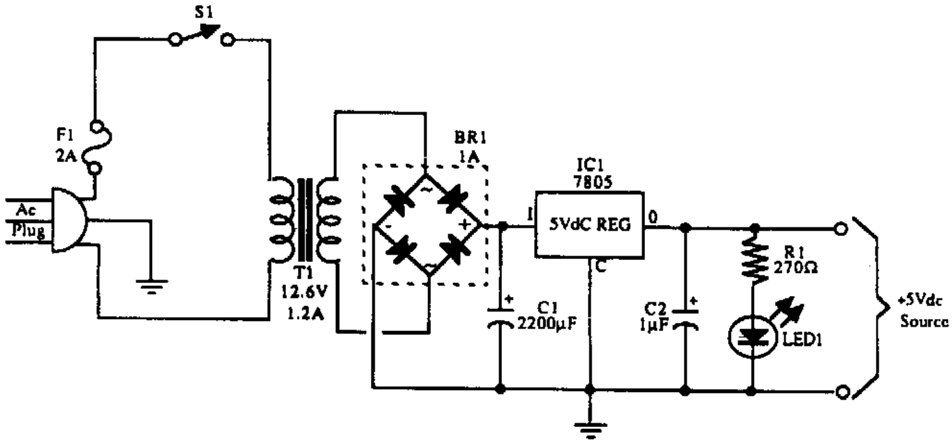
+5-V AT 1-A SUPPLY WITH +3- TO +5-V INPUT



MAXIM *Fig. 68-30*

A MAX741U switching-mode power-supply controller and a switching FET Q1 are used to provide +5 V at 1 A.

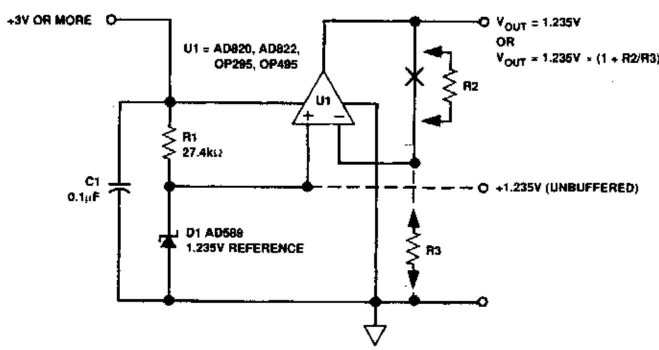
5-Vdc REGULATED SUPPLY



MCGRAW-HILL

Fig. 68-31

BUFFERED REFERENCE SUPPLY

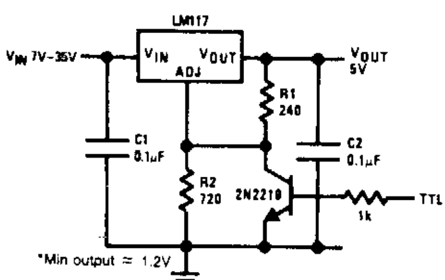


ANALOG DEVICES

Fig. 68-32

This buffered reference (for 1.23 V or more) uses a supply voltage of greater than 3 V.

5-V LOGIC REGULATOR WITH ELECTRONIC SHUTDOWN

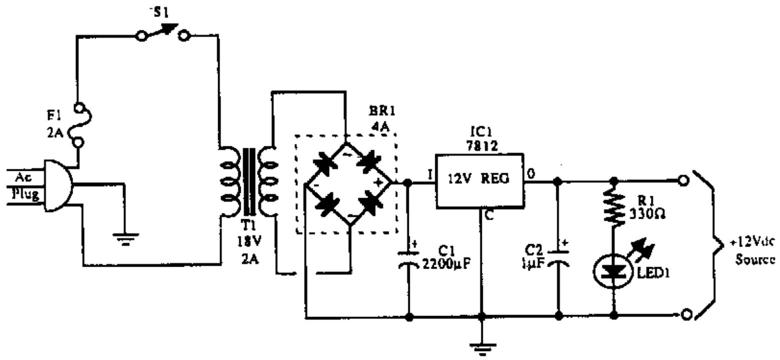


The circuit will shut down to 1.2 V under fault conditions.

NATIONAL SEMICONDUCTOR

Fig. 68-33

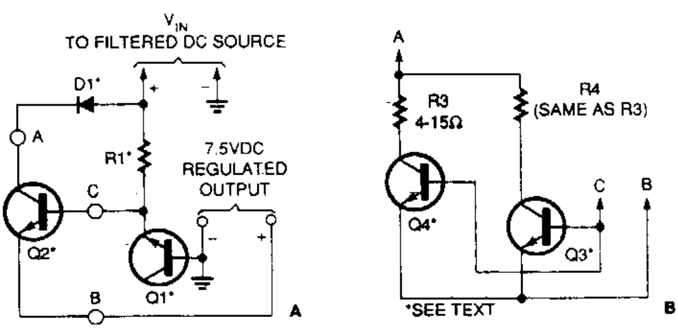
12-Vdc REGULATED SUPPLY



MCGRAW-HILL

Fig. 68-34

JUNKED TRANSISTOR REGULATORS

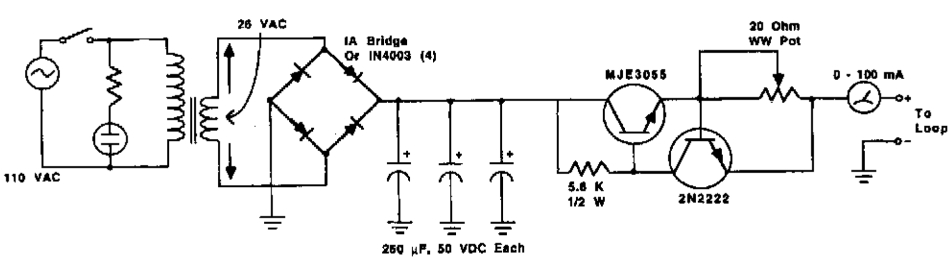


POPULAR ELECTRONICS

Fig. 68-35

Old transistors can make excellent regulators. Simply use one as a Zener to control the base current to another transistor (Fig. 68-35A). If the pass transistor cannot supply enough current, you can use two pass transistors in its place (Fig. 68-35B).

TELEPRINTER LOOP SUPPLY

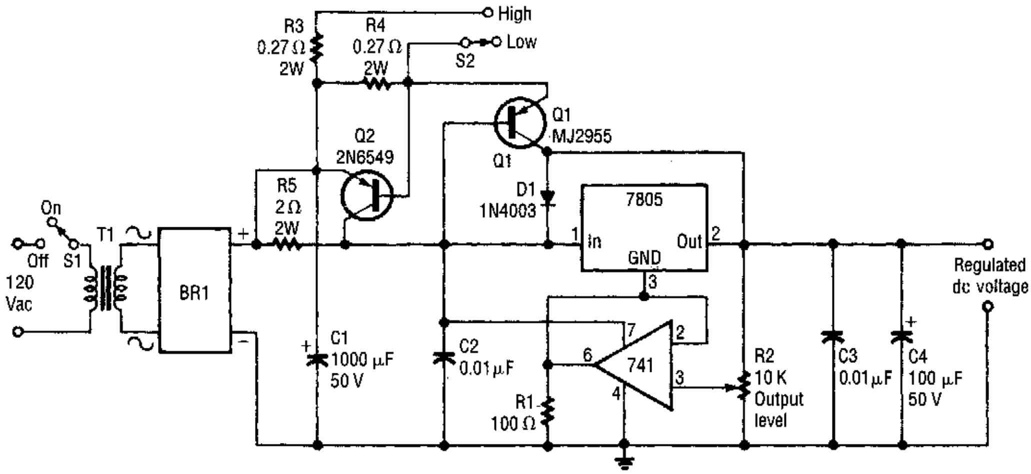


73 AMATEUR RADIO TODAY

Fig. 68-36

A circuit to power a teleprinter, using transistors as current-controlling devices. The power supply used provides a constant current in a loop, normally 60 mA or 20 mA, depending on the machine.

5-A CONSTANT-VOLTAGE SUPPLY



ELECTRONICS NOW *Fig. 68-37*

This constant-voltage supply has a variable output. It can supply more than 5 A, and has two switchable current limits.

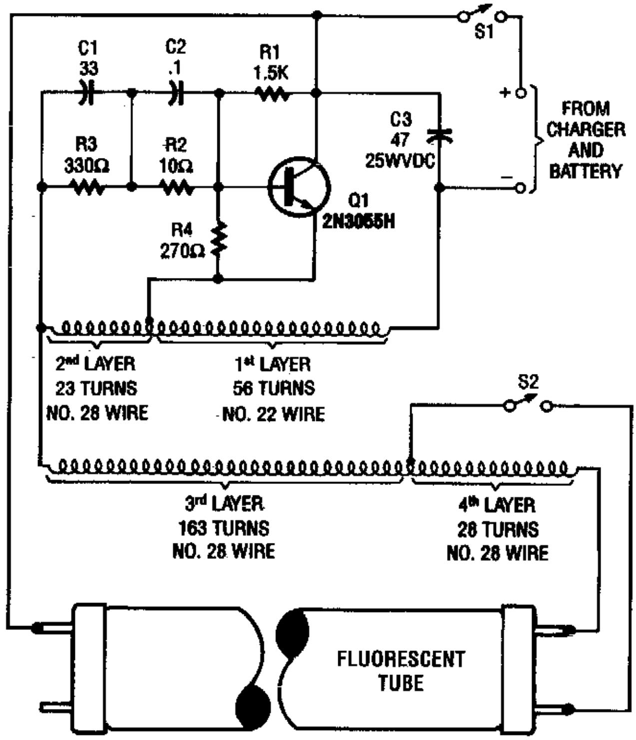
69

Power Supply Circuits (High Voltage)

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

Fluorescent Lamp 12-V Supply
High-Voltage Regulator
Night-Vision Scope Power Supply
High-Voltage Power-Supply Control Circuit
–100-Vdc Supply
ac-Operated He-Ne Power Supply
HV Regulator with Foldback Current Limit
Kirlian Device Supply
High-Voltage Tripler
200-V Regulator
Pulse-Width Modulated Laser Supply

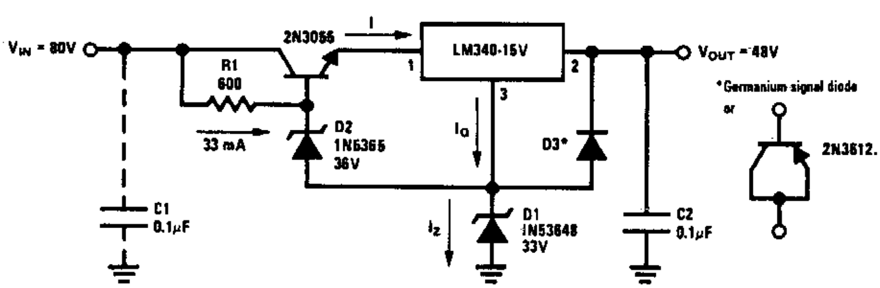
FLUORESCENT LAMP 12-V SUPPLY



POPULAR ELECTRONICS *Fig. 69-1*

This high-voltage power supply can operate fluorescent tubes from a 12-V source, even if the tube has a defective filament. It essentially is an oscillator that excites a home-made autotransformer. T1 is wound on a ferrite rod 5/16" diameter by 1 7/8" long, in layers. S2 is an optional lamp filament switch.

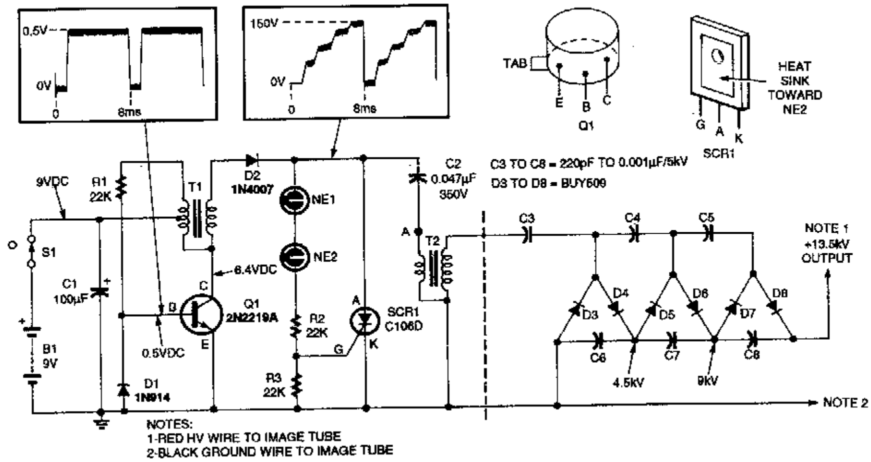
HIGH-VOLTAGE REGULATOR



NATIONAL SEMICONDUCTOR *Fig. 69-2*

This circuit produces 48 V from an 80-V input.

NIGHT-VISION SCOPE POWER SUPPLY



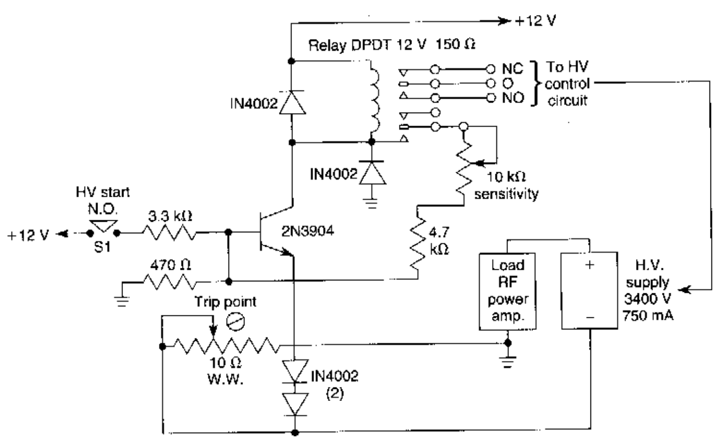
ELECTRONICS NOW

Fig. 69-3

This high-voltage power supply has an inverter around Q1 that supplies 150-V pulses to the converter of SCR1 and C2. The output of T2 is a 4.5-kV pulse that is multiplied by the voltage-tripler network (right) to produce 13.5 kV.

T1 is a 3-kΩ to 500-Ω CT transistor audio transformer, T2 is a flash tube trigger transformer with a 6-kV secondary.

HIGH-VOLTAGE POWER-SUPPLY CONTROL CIRCUIT

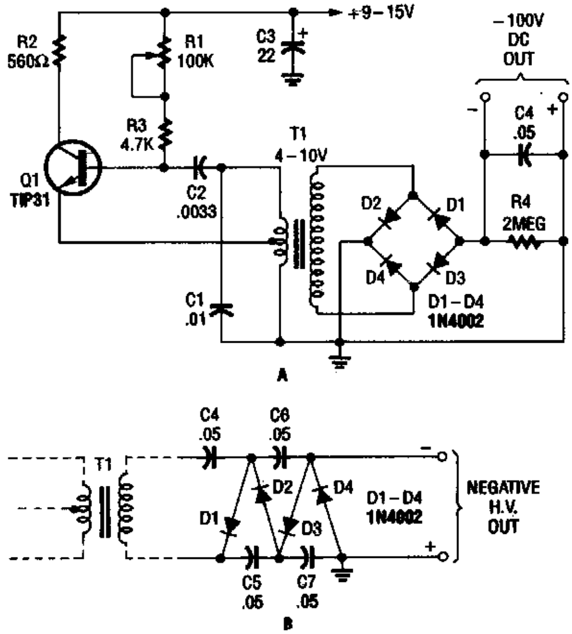


WILLIAM SHEETS

Fig. 69-4

To start the HV supply, S1 is pressed, latching the relay. The 10-kΩ pot is set so that the relay just latches. When HV current becomes excessive (arc-over, etc.), an excessive voltage is developed across the 10-Ω WW pot, cutting off the 2N3904, causing the relay to unlatch. This circuit was used in a 1000-W linear RF power amplifier.

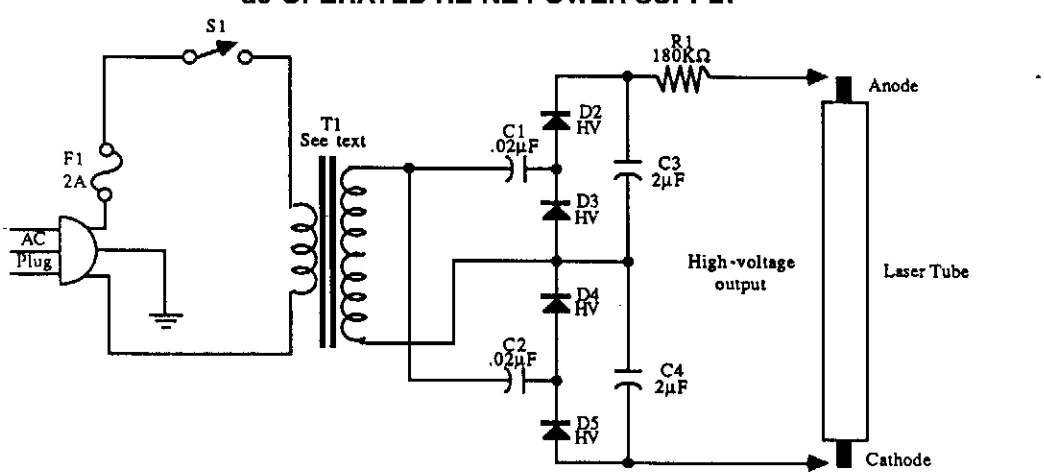
–100-Vdc SUPPLY



POPULAR ELECTRONICS *Fig. 69-5*

The combination Hartley oscillator/step-up transformer shown in A can generate significant negative high-voltage—especially if the voltage output of the transformer is multiplied by the circuit in Fig. 69-5B. T1 is a small low-voltage filament transformer of around 4- to 10-Vac output, 120-V primary.

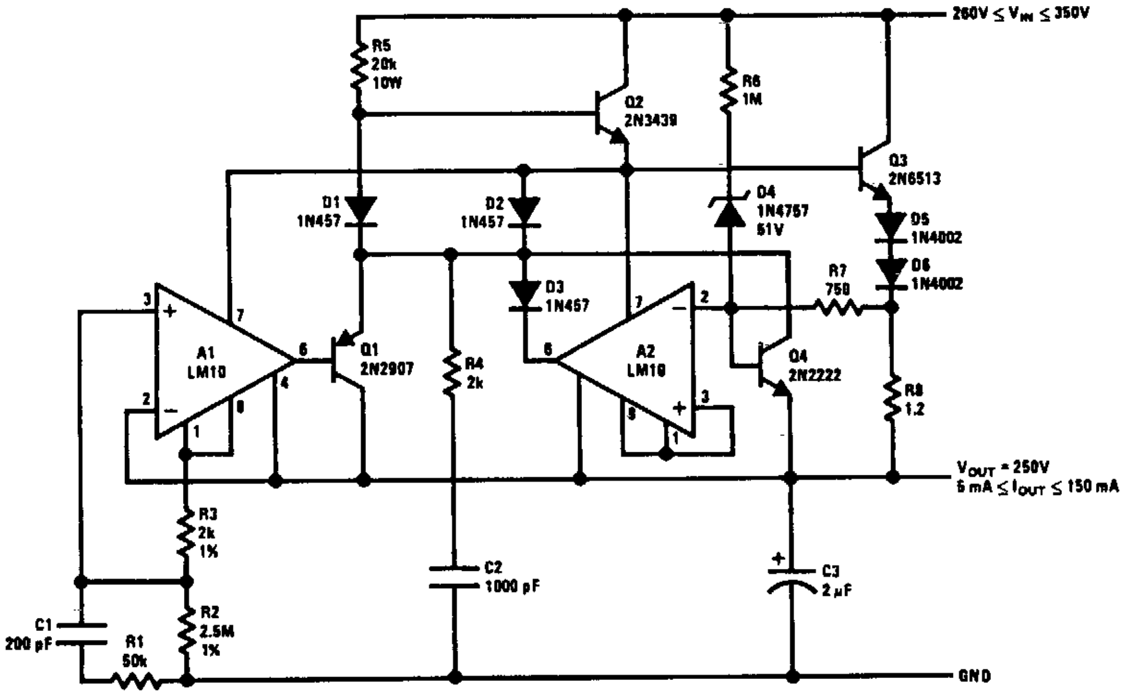
ac-OPERATED HE-NE POWER SUPPLY



McGRAW-HILL *Fig. 69-6*

T1 is a 120-V to 1000-V step-up 60-Hz transformer. C1, C2, C3, C4 and D2 through D5 form a voltage quadrupler. The initial voltage is 4 to 5 kV, which drops when the laser tube fires.

HV REGULATOR WITH FOLDBACK CURRENT LIMIT



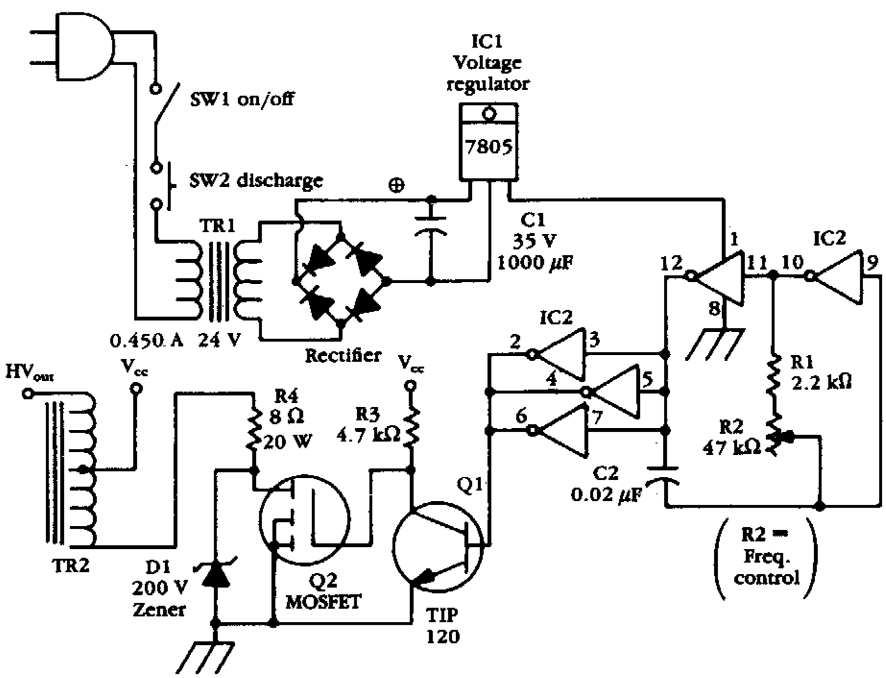
NATIONAL SEMICONDUCTOR

Fig. 69-7

The output current is sensed across R8. This is delivered to the current-limit amplifier through R7, across which the foldback potential is developed by R6 with a threshold determined by D4. The values given limit the peak power below 20 W and shut off the pass transistors when the voltage across them exceeds 310 V. With unregulated input voltages above this value, start-up is initiated solely by the current through R5. Q4 is added to provide some control on current before A2 has time to react.

The circuit is stable with an output capacitor greater than about 2 μF. Spurious oscillations in current limit are suppressed by C2 and R4, while a strange, latch-mode oscillation coming out of current limit is killed with C1 and R1.

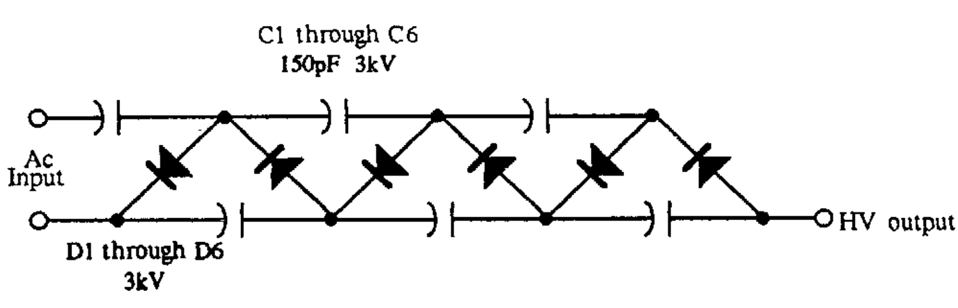
KIRLIAN DEVICE SUPPLY



McGRAW-HILL **Fig. 69-8**

This device is essentially a high-voltage variable-frequency ac supply. A CD4049 IC multivibrator circuit drives a Darlington connected transistor pair, which drives TR2, an HV transformer.

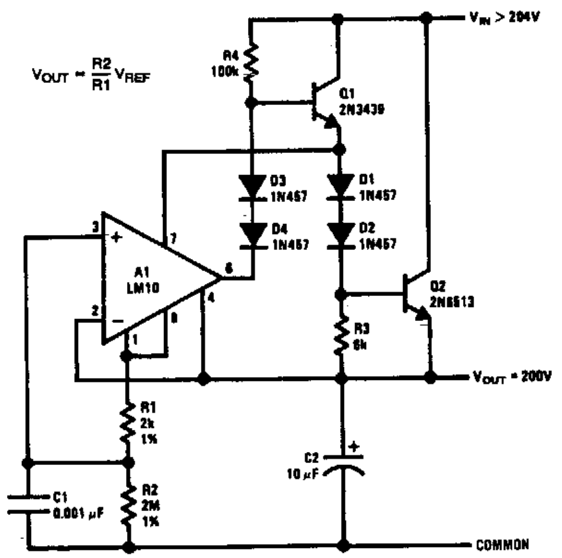
HIGH-VOLTAGE TRIPLER



McGRAW-HILL **Fig. 69-9**

This tripler is useful for low-current and high-voltage applications. The capacitors can be 0.001-μF, 3- to 6-kV discs, and the diode's 3-kV units, or three each IN4007 in series.

200-V REGULATOR



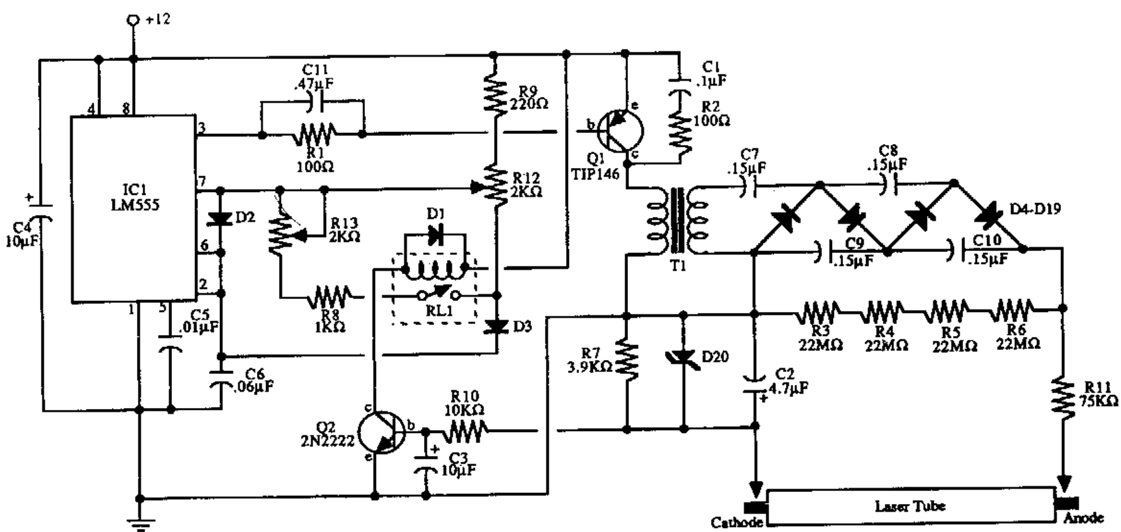
NATIONAL SEMICONDUCTOR

Fig. 69-10

With high-voltage regulators, powering on the IC through the drive resistor for the pass transistors can become quite inefficient. This is avoided with the circuit shown. The supply current for the IC is derived from Q1. This allows R4 to be increased by an order of magnitude without affecting the dropout voltage.

Selection of the output transistors will depend on voltage requirements. For output voltages above 200 V, it might be more economical to cascade lower-voltage transistors.

PULSE-WIDTH MODULATED LASER SUPPLY



McGRAW-HILL

Fig. 69-11

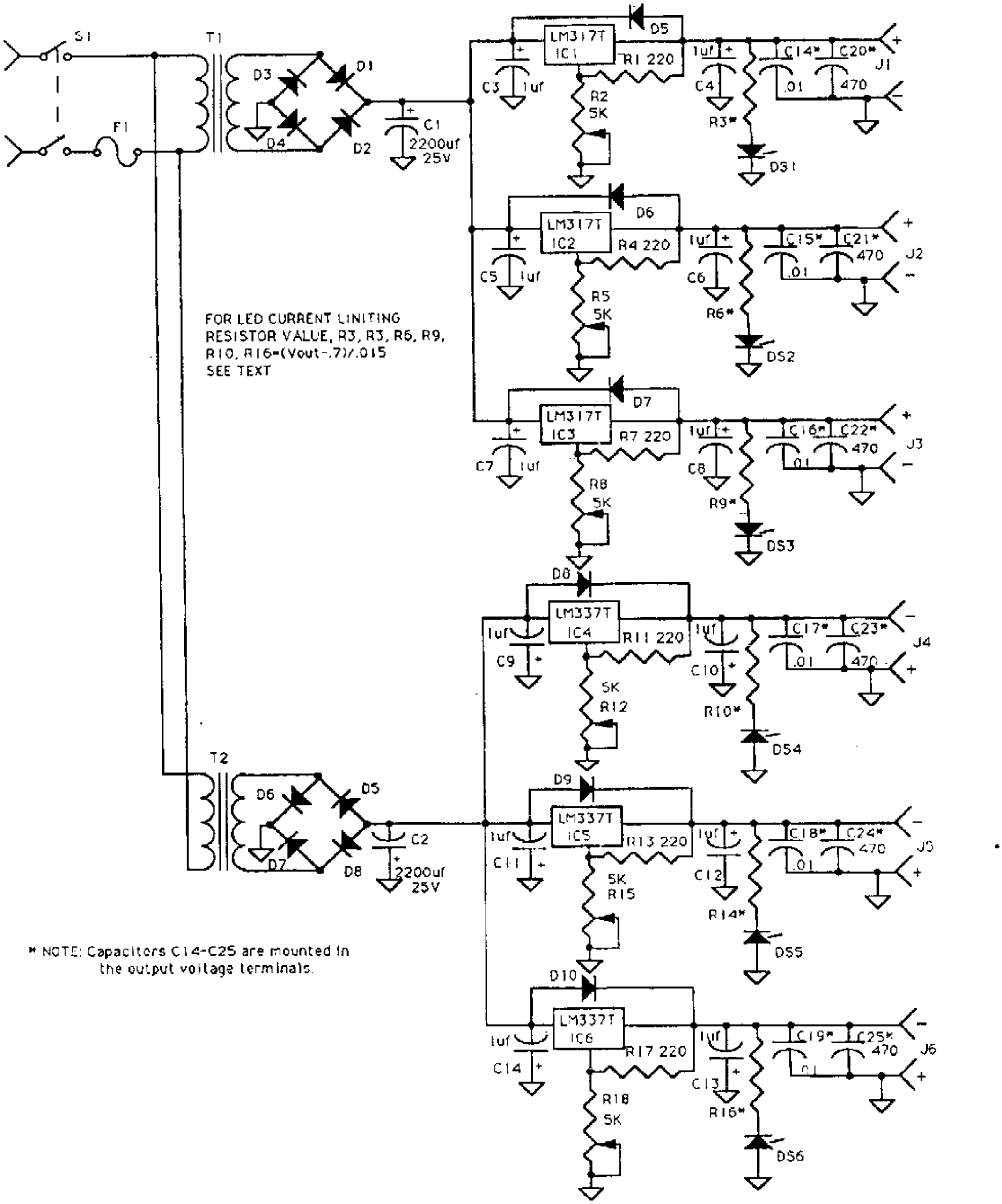
IC1 initially provides drive for Q1 and HV transformer T1, and it rectifies D4 through D19. When the laser tube ignites, Q2 is triggered; this activates relay RL1, reducing the duty cycle. R13 controls the duty cycle of the pulses through the laser tube.

70

Power Supply Circuits (Multiple Output)

The sources of the following circuits are contained in the Sources section, which begins on page 706. The figure number in the box of each circuit correlates to the entry in the Sources section.

EXPERIMENTER'S POWER SUPPLY

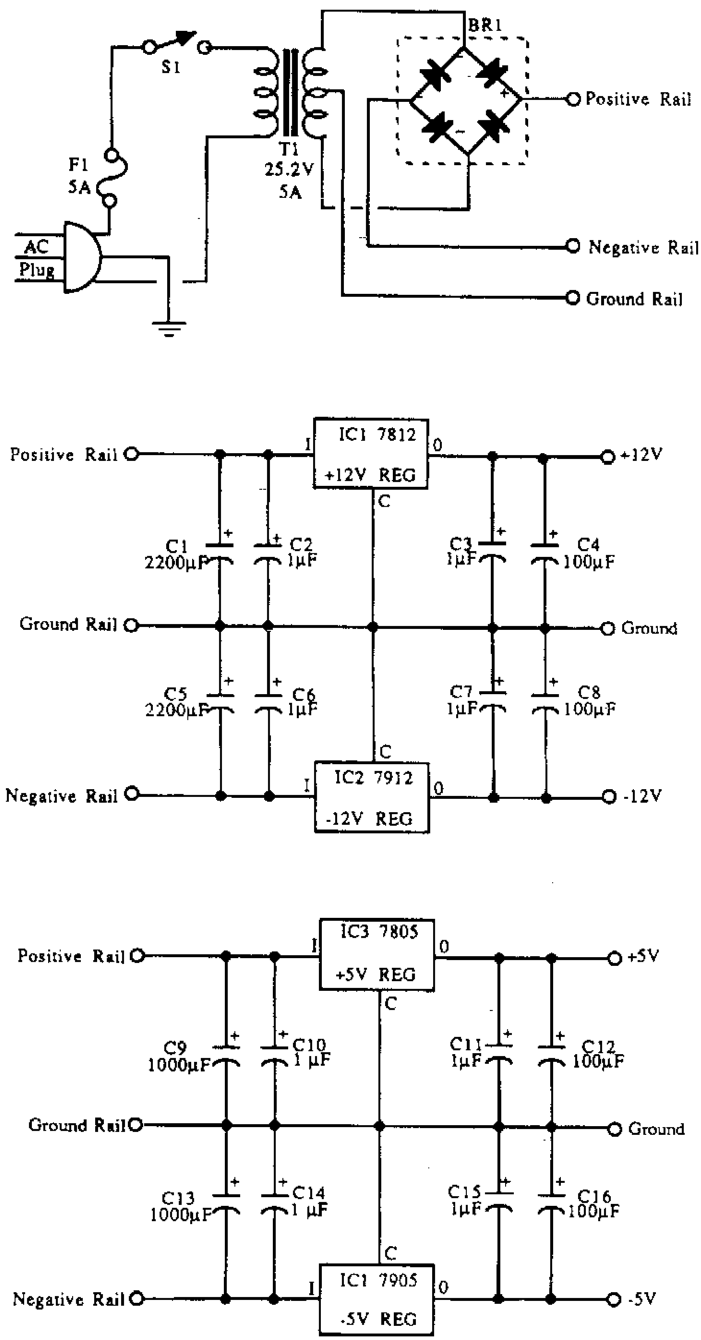


73 AMATEUR RADIO TODAY

Fig. 70-1

Passive linear IC regulators are used to make up a supply delivering +12, +9, +5, –5, –9, and –12 Vdc. T1 and T2 are 12-V, 3-A transformers.

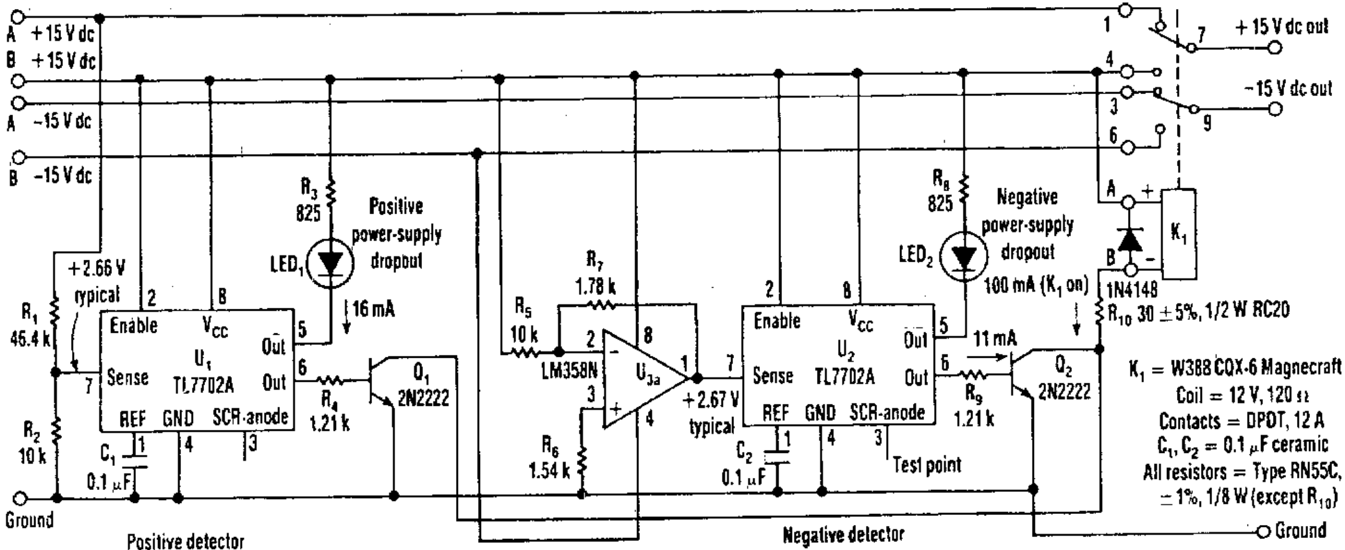
QUAD POWER SUPPLY



McGRAW-HILL

Fig. 70-2

ACTIVATE BACK-UP POWER SUPPLY



ELECTRONIC DESIGN

Fig. 70-3

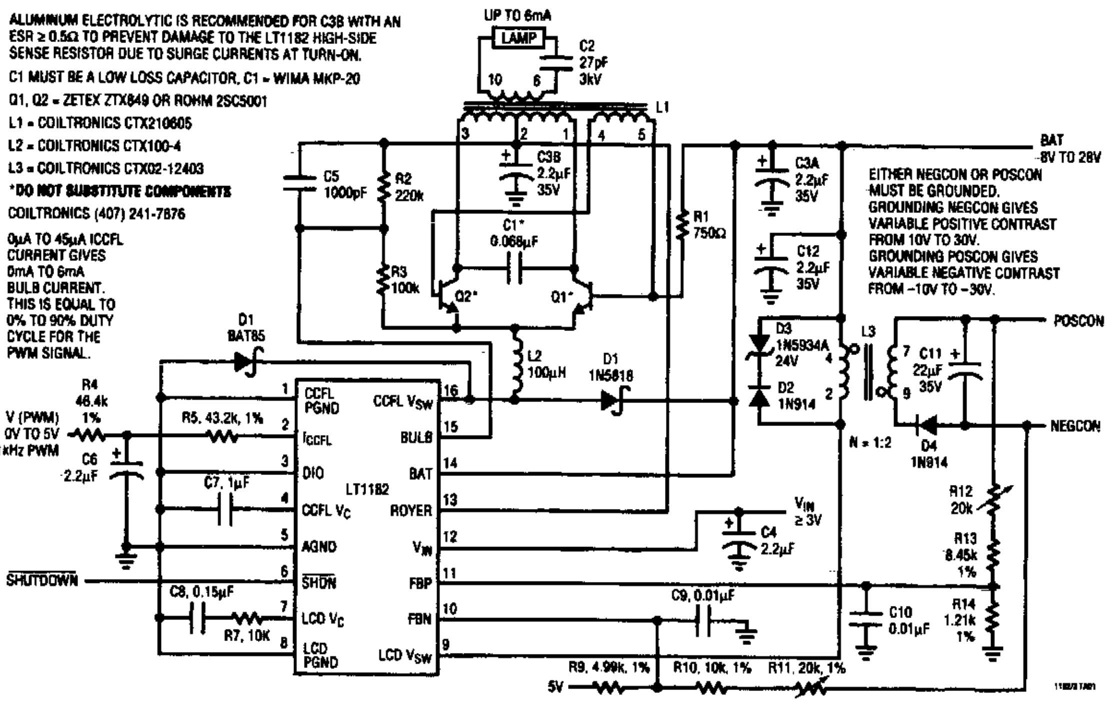
A circuit, which can be built around two TI TL7702 chips, monitors a 15-V power supply and activates a relay to turn on a back-up supply if the voltage drops below ±14.1 V. With the back-up in place, the malfunctioning supply can be replaced without any down time. The TL7702 power-supply voltage supervisor chips are intended for use as reset controllers in microprocessor systems, but they work well in their modified form for this application.

One chip detects the positive supply (U1), and the other is used for the negative (U2). A pair of output-relay drive transistors, Q1 and Q2, form a wired OR circuit so that relay K1 is activated by the positive or negative voltage detector that switches U1 or U2 on.

The supervisor chips have a direct connection to the input comparator so that the trigger level is set by a resistor-divider network (R1 and R2) at the sense-input pin. These chips also have an internal, stable, reference-voltage source set at +2.53 V, typical. The positive-sensed voltage drops enough to activate the comparator, its output goes low, switches the internal gate, and triggers the silicon-controlled rectifier (SCR). The output comparator then forces the two output transistors to switch, one high and one low. The output transistor (pin 5) turns on the light-emitting diode (LED), and the output transistor (pin 6) turns on the relay driver, Q1.

The negative detector is preceded by half of the dual op amp LM358N (U3a)—an inverting amplifier with a gain of −0.178. R5 connects to the −15 Vdc being sensed. The output of U3a is usually set at 2.67 V, higher than the +2.53-V reference voltage. Therefore, no switching occurs. If the −15-Vdc voltage decreases, U2 switches Q2 and activates K1 in the same manner as described for the U1-Q1 positive detector. R3, R4, and R8 through R10 serve as current-limiting resistors.

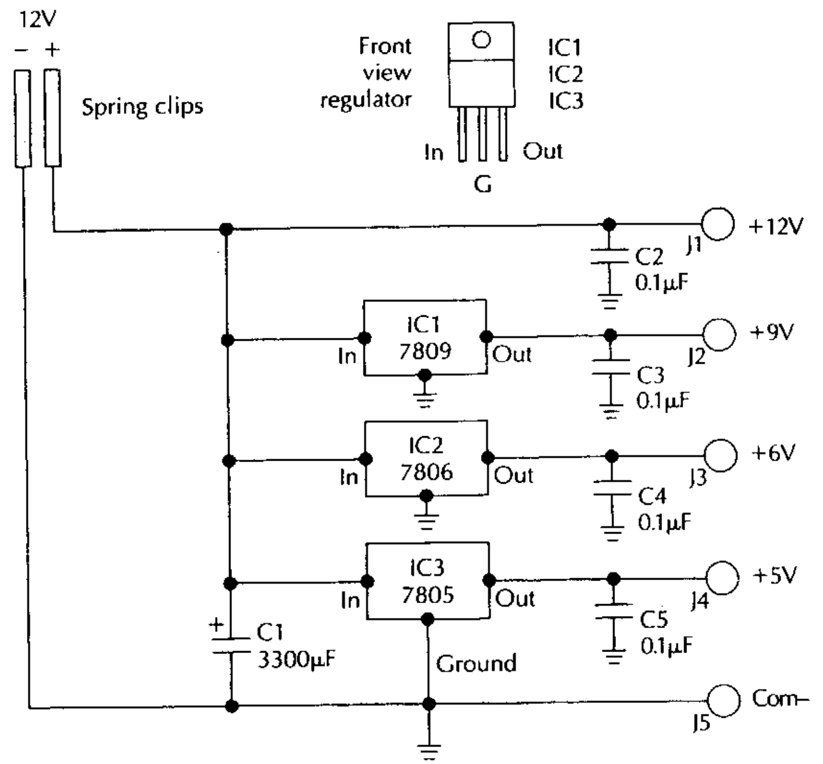
CCFL SUPPLY WITH VARIABLE CONTRAST



LINEAR TECHNOLOGY *Fig. 70-4*

The figure is a complete floating CCFL circuit with variable negative/variable positive-contrast voltage capability, based on the LT1182. Lamp current is programmable from 0 mA to 6 mA using a 0- to 5-V 1-kHz PWM signal at 0% to 90% duty cycle. LCD contrast output voltage polarity is determined by which side of the transformer secondary (either POSCON or NEGCON) the output connector grounds. In either case, LCD contrast output voltage is variable from an absolute value of 10 V to 30 V. The input supply voltage range is 8 V to 28 V. The CCFL converter is optimized for photometric output per watt of input power. CCFL electrical efficiency up to 90% is possible and requires strict attention to detail. LCD contrast efficiency is 82% at full power.

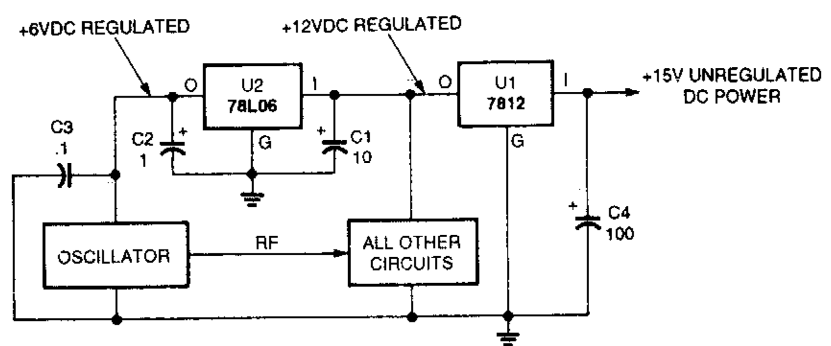
dc POWER SOURCE FOR EXPERIMENTS



McGRAW-HILL *Fig. 70-5*

This supply uses IC regulators to supply +5, +6, +9, and +12 volts regulated from a nominal 12-V supply.

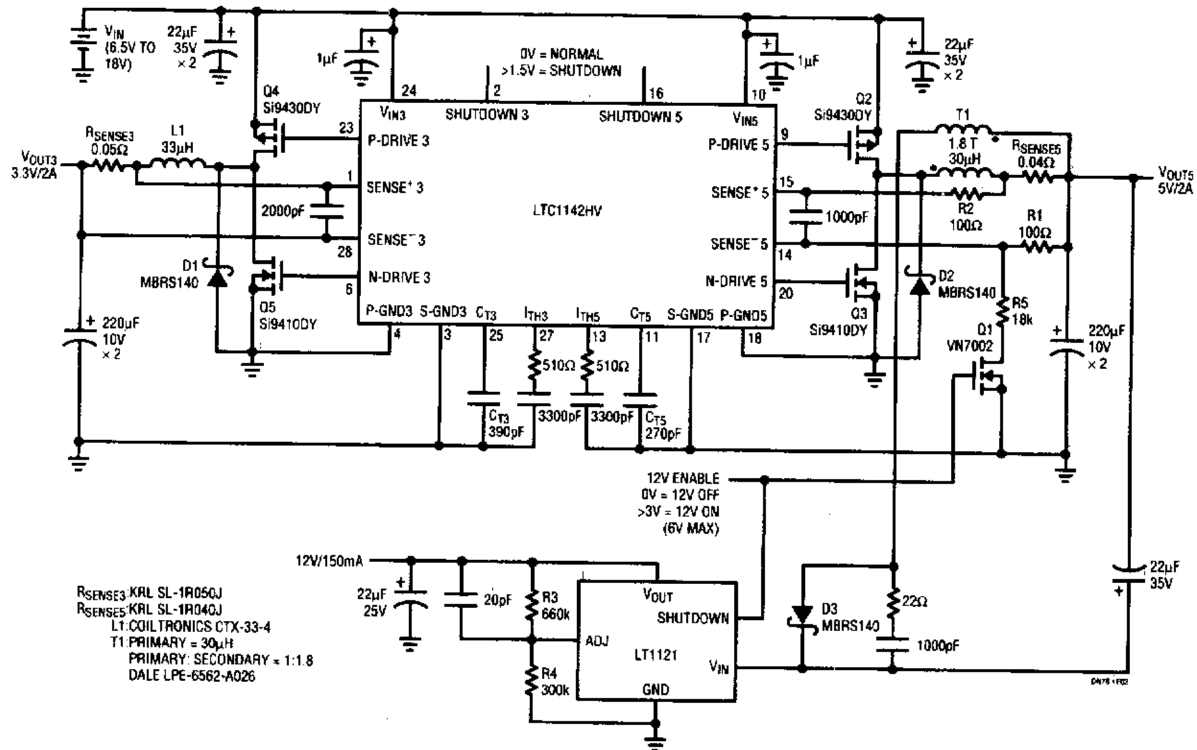
STABLE VFO POWER SUPPLY



POPULAR ELECTRONICS *Fig. 70-6*

A dc power-distribution system for a stable oscillator should use a separate voltage regulator just for the oscillator circuit.

HIGH-EFFICIENCY TRIPLE-OUTPUT SUPPLY FOR NOTEBOOK COMPUTERS



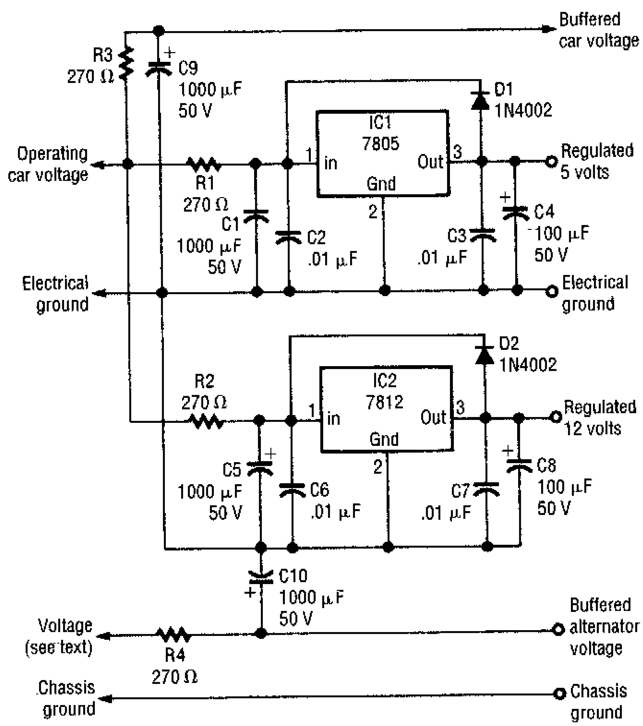
LINEAR TECHNOLOGY

Fig. 70-7

The circuit is configured to provide output voltages of 3.3 V, 5 V, and 12 V. The current capability of both the 3.3-V and 5-V outputs is 2 A (2.5 A peak). The logic controlled 12-V output can provide 150 mA (200 mA peak), which is ideal for flash memory applications. The operating efficiency shown in the figure exceeds 90% for both the 3.3-V and 5-V sections.

The 3.3-V section for the circuit in the figure is comprised of the main switch Q4, synchronous switch Q5, inductor L1, and current shunt R_{SENSE}. Current-sense resistor R_{SENSE} monitors the inductor current and is used to set the output current according to the formula $OUT = 100\ mV/R_{SENSE}$. Advantages of current control include excellent line and load transient rejection, inherent short-circuit protection, and controlled start-up currents. Peak inductor currents for L1 and T1 for the circuit in the figure are limited to $150\ mV/R_{SENSE}$ or 3.0 A and 3.75 A, respectively.

GENERAL-PURPOSE POWER SUPPLY FOR AUTOMOTIVE PROJECTS

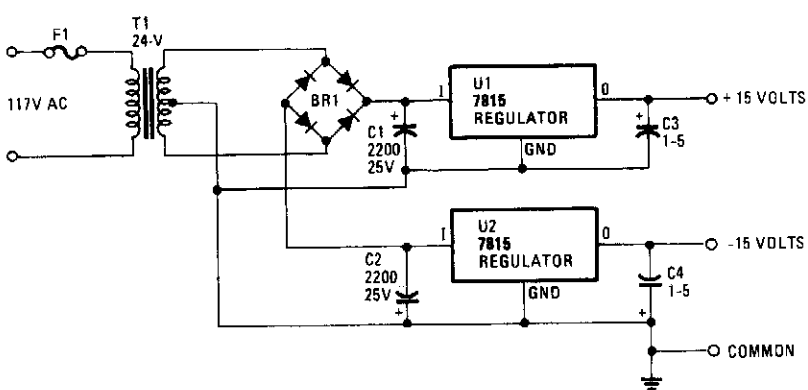


ELECTRONICS NOW

Fig. 70-8

This supply produces 12 V and 5 V for a variety of automotive projects. F4 is connected directly to the alternator field winding (usable only if your car has a separate regulator).

±15-V POWER SUPPLY



RADIO-ELECTRONICS

Fig. 70-9

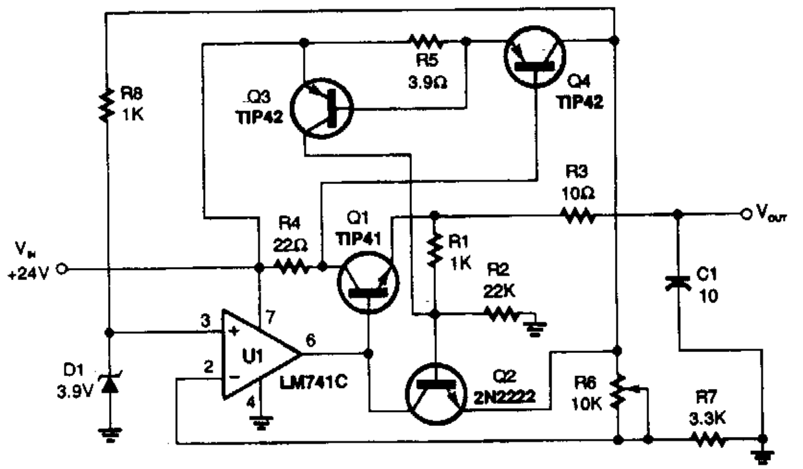
A simple bridge rectifier feeds two IC regulators. This circuit should be useful for op-amp circuitry.

71

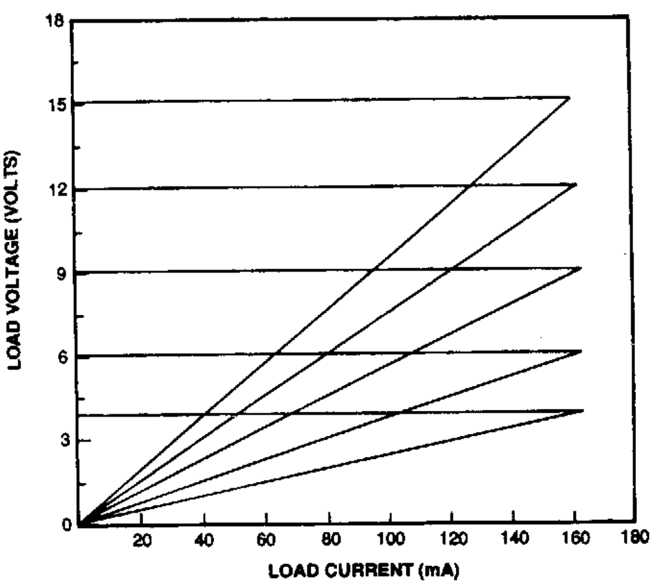
Power Supply Circuits (Variable Output)

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

VARIABLE VOLTAGE REGULATOR WITH CURRENT CROWBAR LIMITING



A



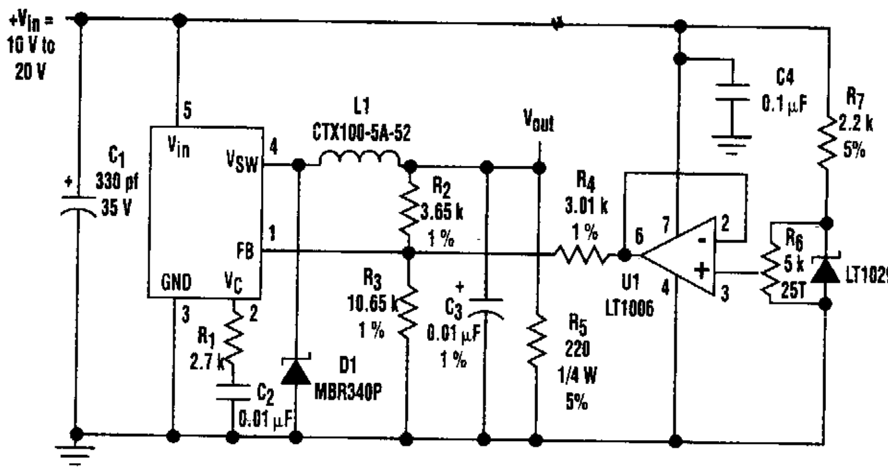
B

POPULAR ELECTRONICS

Fig. 71-1

The variable voltage regulator with current-crowbar limiting, shown in Fig. 71-1A, overcomes the disadvantages of constant and foldback limiting. As you can see in the graph (Fig. 71-1B), the current crowbar quickly shuts down the supplied power when a preset current is exceeded. It also has excellent load regulation over its operating range.

ADJUSTABLE 0- TO 5-V SUPPLY



ELECTRONIC DESIGN

Fig. 71-2

Although linear-regulator ICs are frequently used in variable power-supply applications, they might not always be the best choice. At low output voltages, power losses in these regulators can cause headaches for designers. For example, if an output current of 1.25 A is required at 1.25 V from an input of 8 V. a regulator (such as the LT317) dissipates more than 10 W.

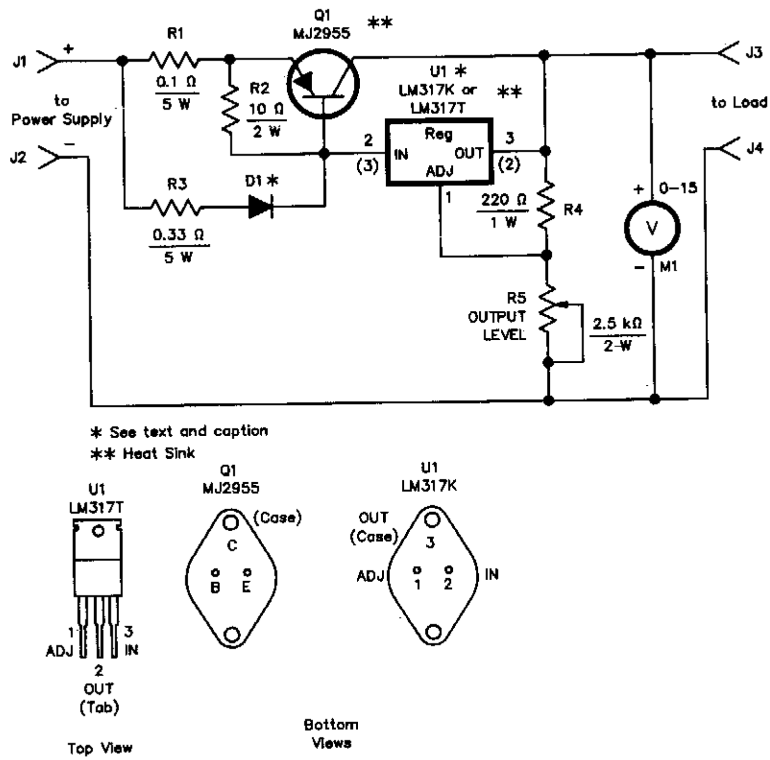
The figure depicts a dc-dc converter that functionally replaces a linear regulator in the just described application. The converter not only eliminates the problem of power loss, but it can be adjusted for output voltages (as low as 25 mV) while delivering an output current of 1.5 A.

The circuit uses a basic positive-buck topology with one exception. A control voltage is applied through R4 to the feedback summing node at pin 1 of the LT1076 regulator IC, making it possible to adjust the output from 0 V to approximately 6 V. This range encompasses the 3.3-V and 5-V logic supply voltages for portable and desktop equipment, as well as battery-pack combinations of one to four cells.

As R4 is driven from 0 to 5 V by the buffer (U1), more or less current is required from R2 to satisfy the loop's desire to hold the feedback summing point at 2.37 V. This forces the converter's output to swing over the range of 0 to 6 V.

The LT1076 is capable of 1.75-A guaranteed output current in this application, and 2 A is typical. If more current is required, the LT1074 can be substituted for the LT1076.

TRANSCEIVER POWER SUPPLY FOR VARIABLE LAB SOURCE

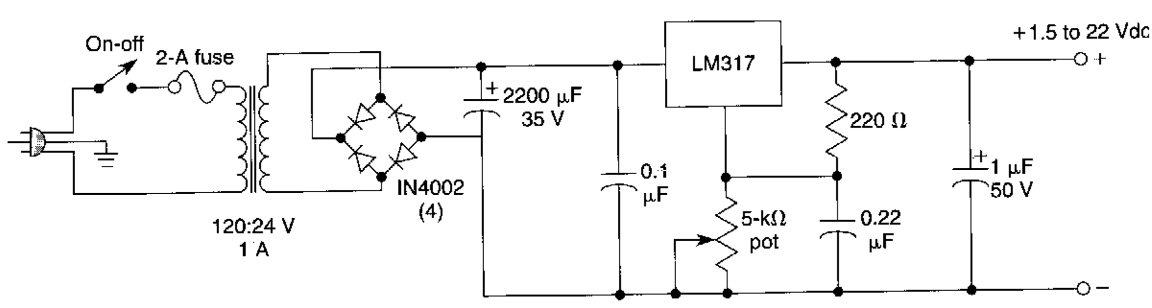


QST

Fig. 71-3

A variable voltage regulator provides 1 to 11 Vdc for lab bench work, using an existing 13.8-V transceiver supply.

ADJUSTABLE POWER SUPPLY

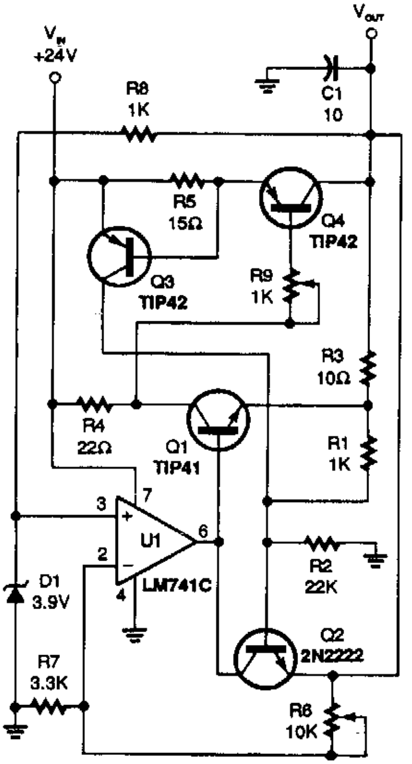


WILLIAM SHEETS

Fig. 71-4

Using an LM317, this supply delivers 1.25 to 22 Vdc for various purposes. The LM317 should be heatsinked. This supply will deliver 600-mA output current.

VARIABLE-VOLTAGE REGULATOR WITH WIDE-RANGE CURRENT LIMITING



POPULAR ELECTRONICS

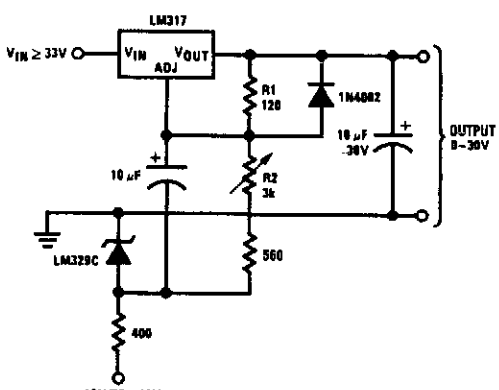
Fig. 71-5

In this circuit, R9 acts as a control to set current limiting. If $R_9 = 0$, limiting occurs at 47 mA. Input is 24 V, output is

$$\frac{(R_6 + R_7)}{R_7} \times \tag{3.9}$$

depending on the setting of R6.

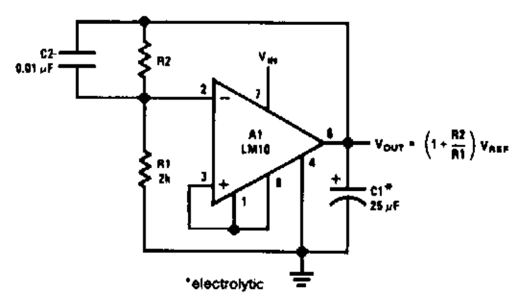
GENERAL-PURPOSE 0- TO 30-V POWER SUPPLY



NATIONAL SEMICONDUCTOR

Fig. 71-6

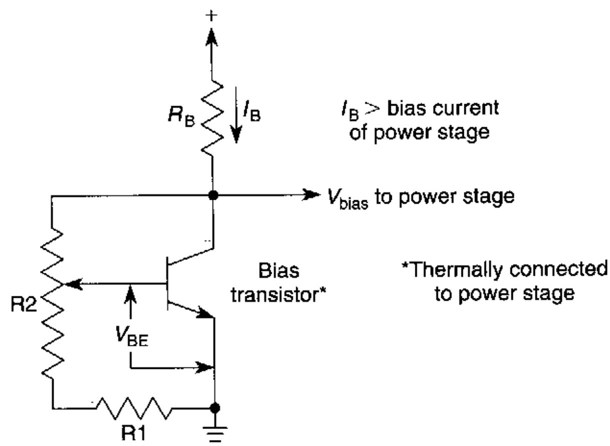
ADJUSTABLE POSITIVE REGULATOR



NATIONAL SEMICONDUCTOR

Fig. 71-7

ADJUSTABLE BIAS REGULATOR



WILLIAM SHEETS

Fig. 71-8

If the wiper of R2 is set at N% rotation, the bias voltage will be:

$$V_{BIAS} = V_{BE}\left(\frac{R_1 + NR_2}{R_1 + R_2}\right)$$

$$V_{BIAS\ MIN} = V_{BE} \qquad V_{BIAS\ MAX} = \left(\frac{R_1}{R_1 + R_2}\right) \quad V_{BE}$$

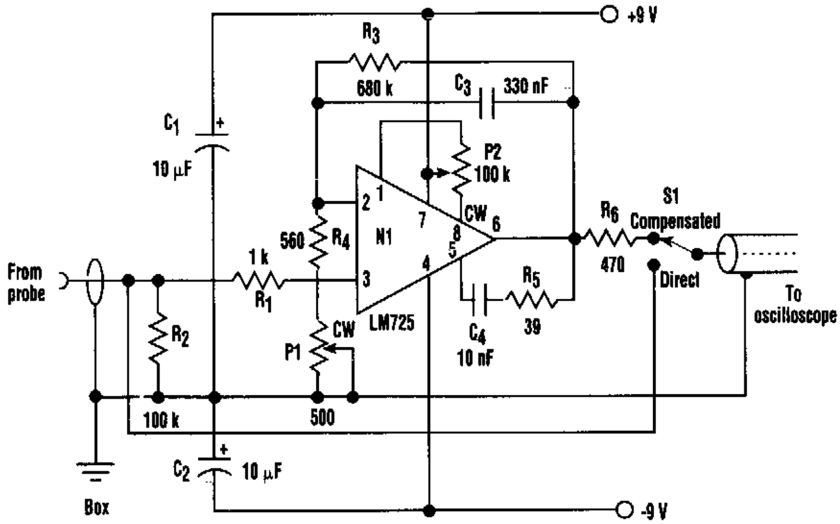
This method derives a bias voltage that tracks V_{BE} of this bias transistor. If the bias transistor is thermally linked to the power stage, tracking over a wide temperature range will result.

72

Probe Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

CURRENT PROBE AMPLIFIER



ELECTRONIC DESIGN

Fig. 72-1

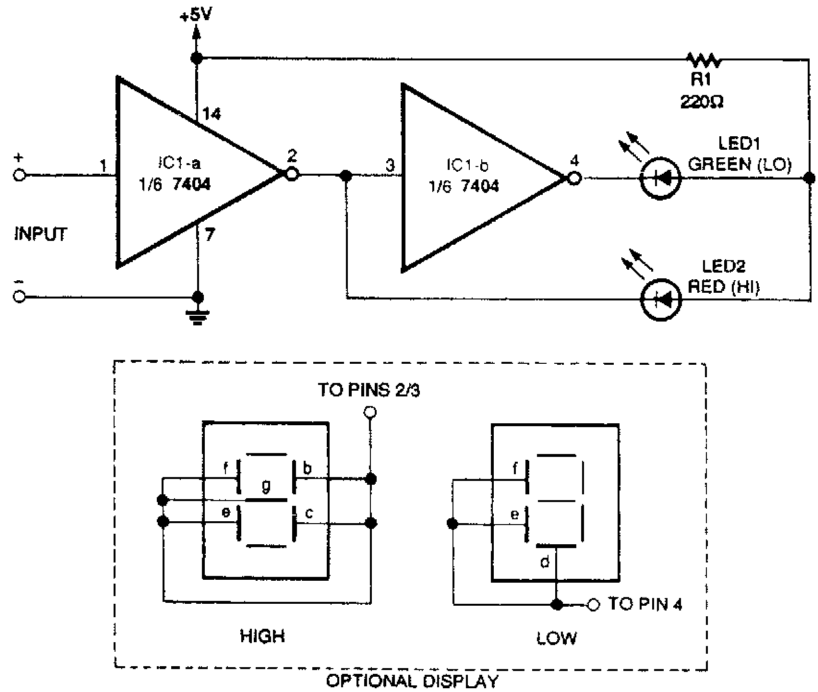
A clamp-on current probe, such as the Tektronix P6021, is a useful means of displaying current waveforms on an oscilloscope. A less-expensive and simple alternative is shown in the figure.

The more sensitive range on the P6021 is 2 mA/mV, but it has a roll-off of 6-dB per octave below 450 Hz. The purpose of the compensator is to counteract the low-frequency attenuation, which is achieved by means of C3 and R4 + P1 in the feedback around op amp N1. It's important that the latter is a low-noise type, such as the LM725 shown in the figure. On top of that, it's necessary at some point to limit the increasing gain with decreasing frequency; otherwise, amplifier noise and drive will overcome the signal. The values shown for C_3 and R_3 give a lower limit of less than 1 Hz.

A test square wave of ±1 mA is fed to the current probe so that P1 can be adjusted for minimum droop or overshoot in the output waveform. It's vital that the sliding core on the probe is fully closed. At high frequencies, the response begins to fall off at 100 kHz. Therefore, for most waveforms, switch S1 is moved to "direct," above a fundamental frequency of, for example, 10 kHz.

This circuit's current consumption is quite low, and it can be battery powered. If a mains power supply is built-in, it must be well screened to prevent hum problems.

SIMPLE LOGIC PROBE WITH ALPHANUMERIC DISPLAY

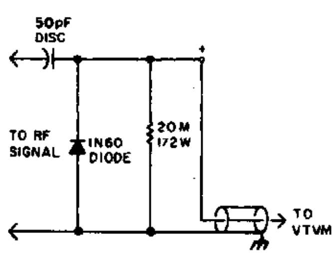


ELECTRONICS NOW

Fig. 72-2

A logic probe also includes BCD decoder module. The red LED lights to indicate a logic high, and the green LED lights to indicate a logic low. This probe circuit will light a green (low) or red (high), and if desired, an alphanumeric display can be obtained with two 7-segment LED displays.

SIMPLE RF PROBE

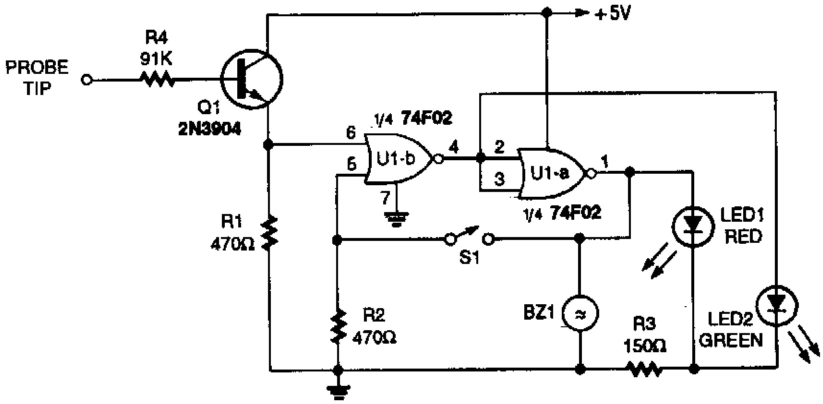


73 AMATEUR RADIO TODAY

Fig. 72-3

Your VTVM can measure peak voltage up to 200 MHz by using this probe. The maximum RF that can be measured is determined by the diode; with a 1N60, the probe is limited to 30 V. To increase the capacity, substitute a higher-voltage small-signal detector diode. House the circuit in a metal enclosure and use shielded wire.

125-MHz LOGIC PROBE

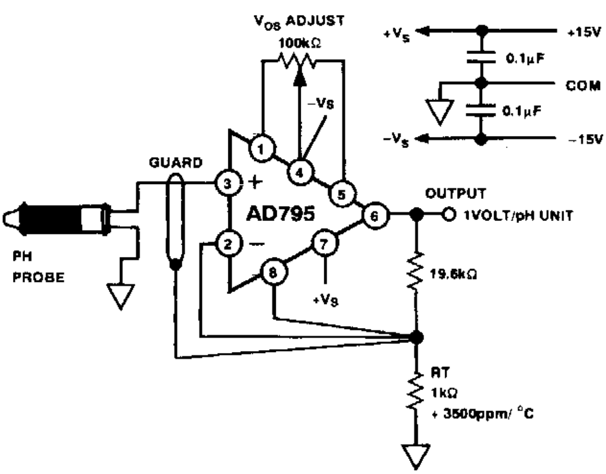


POPULAR ELECTRONICS

Fig. 72-4

This logic probe features either high-low (LED) indication or latching operation. When S1 is closed, the indication of a pulse is latched and the red LED1 stays on. Piezoelectric buzzer BZ1 is used as a beeper to sound that a logic high is preset.

pH PROBE AMPLIFIER

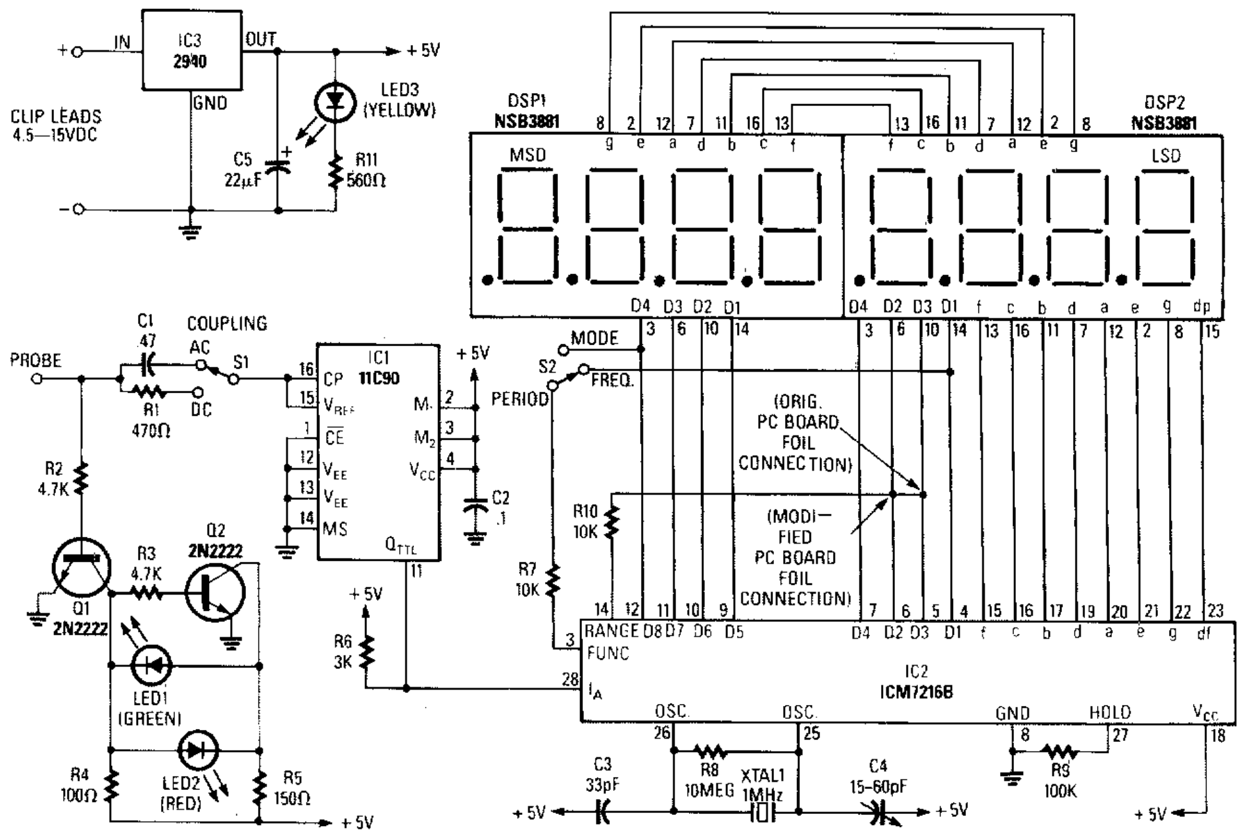


ANALOG DEVICES

Fig. 72-5

The low-noise precision FET op-amp AD795 has around 1014-Ω common-mode impedance, low-offset voltage (250 μV_{max}) and L13 μV_C drift make this device ideal for low-voltage measurements from high-impedance sources.

8-DIGIT 100-MHz FREQUENCY PROBE



RADIO-ELECTRONICS

Fig. 72-6

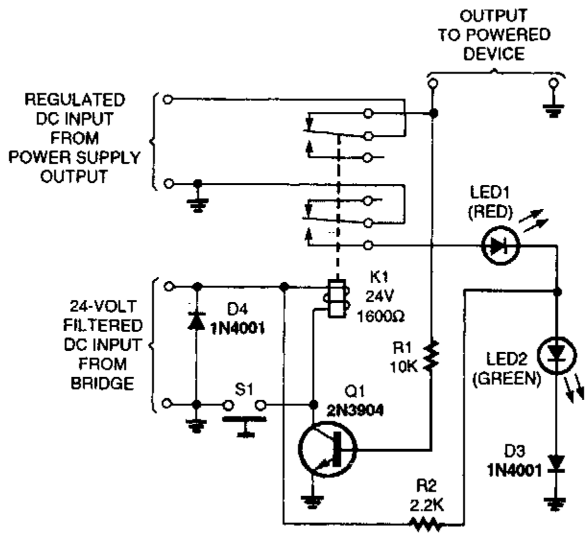
Small enough to mount in a probe, this frequency counter circuit is good to 100 MHz. It operates from +5 to +15 Vdc. An 11C90 prescaler drives a 10-MHz counter chip (ICM7216B). Note the dotted line connecting R10 with pins 5 and 6 of IC2; that variable connection controls the decimal point and total count appearing on DSP1 and DSP2. The relative intensities and durations of ON/OFF time for LED1 (green) and LED2 (red) give a rough indication of logic level and duty cycle.

73

Protection Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SHORT-CIRCUIT PROTECTION CIRCUIT

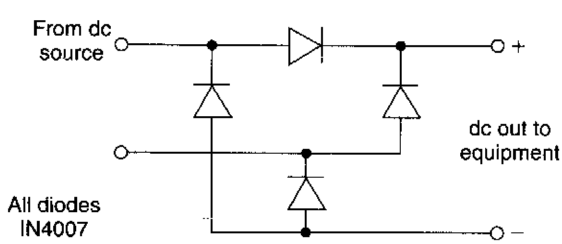


POPULAR ELECTRONICS *Fig. 73-1*

When S1 is pressed, the coil of K1 is energized, closing its normally open contacts. If the regulated dc input is between 1 and 24 V, that voltage feeds the base of Q1 through R1, turning on the transistor, and latching the relay. When that occurs, LED2 glows indicating that all is okay.

If there is a short to ground at the circuit's output (i.e., in the device being powered), the voltage that feeds the base of Q1 goes to zero, turning off the transistor. Then, LED1 glows because K1 is de-energized to indicate the short circuit.

POLARITY PROTECTOR



The use of a four-diode bridge guarantees correct polarity irrespective of input dc polarity. Remember that two diode drops (about 1.2 to 1.5 V) are lost from the input voltage using this circuit.

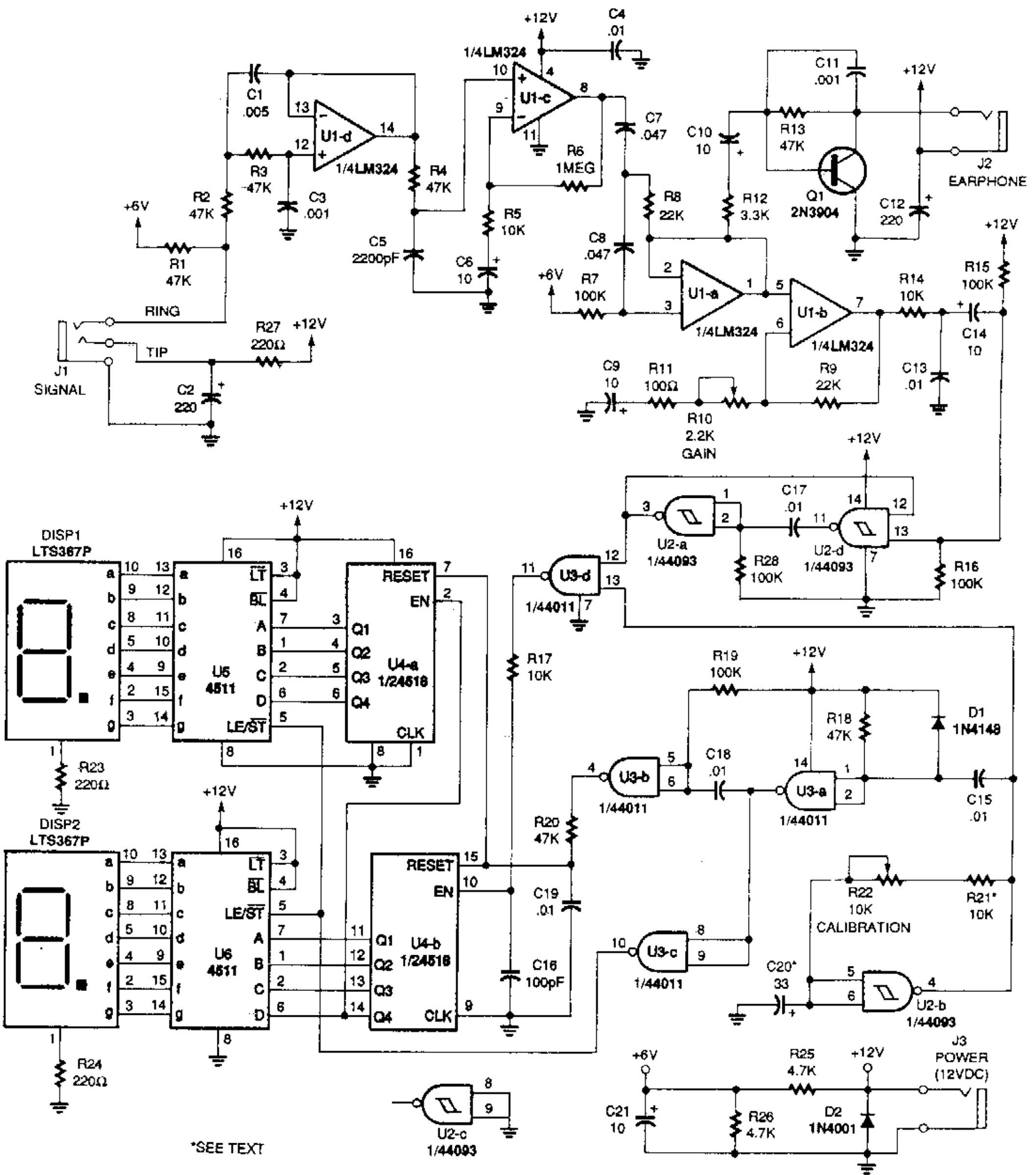
WILLIAM SHEETS *Fig. 73-2*

74

Radar Detector Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

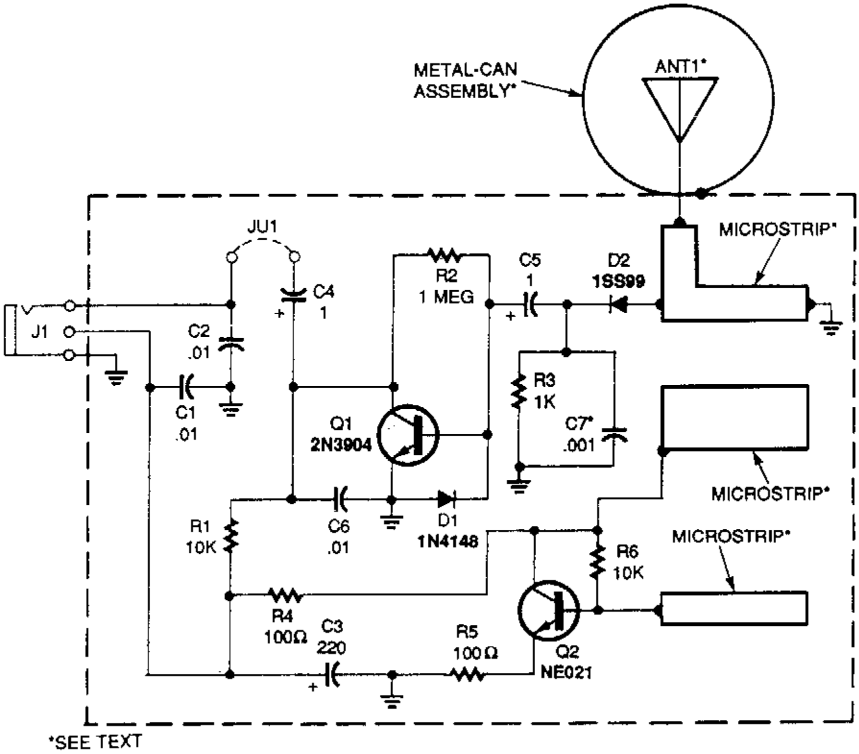
DISPLAY BOARD FOR RADAR GUN



POPULAR ELECTRONICS ***Fig. 74-1***

This circuit takes signal (doppler) from a radar gun, amplifies and limits it, and feeds the frequency into a counter (U4) and display circuit (DISP1, DISP2, U5, U6). Counter calibration is set by clock circuit U2B. Calibration is obtained via R21 and R22. R21 can be changed if kilometers/hour readout is desired.

2.6-GHz OSCILLATOR FOR RADAR SPEED GUN



POPULAR ELECTRONICS *Fig. 74-2*

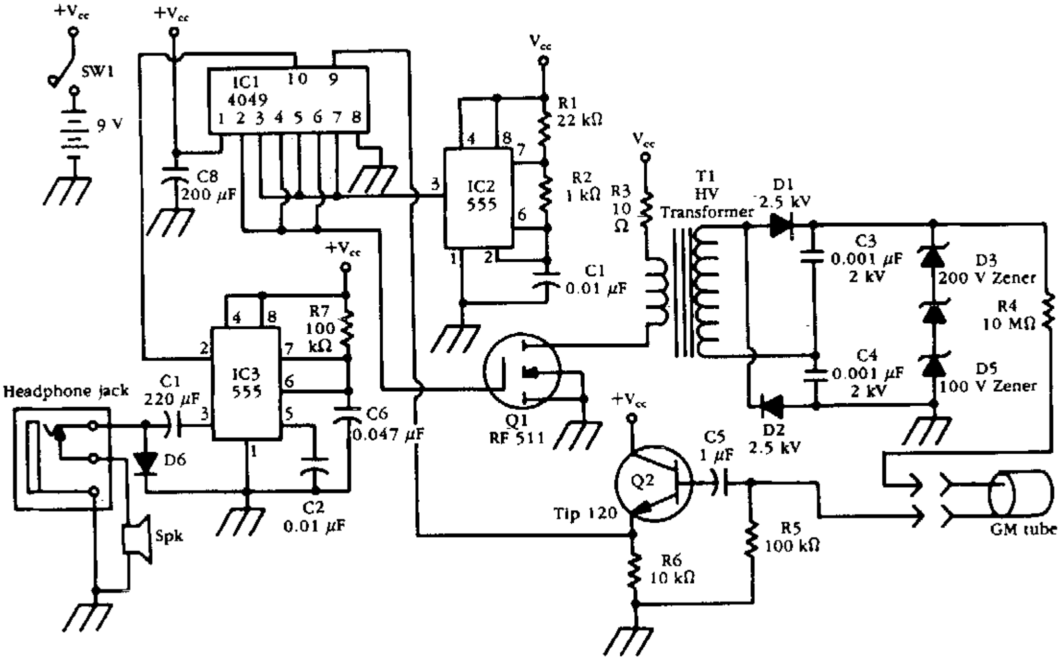
This circuit consists of 2.6-GHz oscillator Q2, a coupling microstripline to ANT1, a 1.1" ¼-wave probe, detector D2, and audio amp Q1. The oscillator feeds power to the antenna, which radiates the signal. The reflected signal from a moving target mixes with the oscillator signal in D2. The resultant beat note (doppler shift) is amplified by Q1 and fed to jack J1, which is used to feed the circuit 12 Vdc.

75

Radiation Detector Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

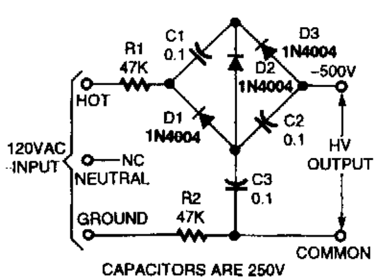
GEIGER COUNTER



McGRAW-HILL *Fig. 75-1*

An HV generator (IC1, IC2, Q1, T1, and associated components) power a G-M tube. A pulse from the GM tube is interfaced through Q2 and IC1 to pulse generator IC3, which drives a speaker.

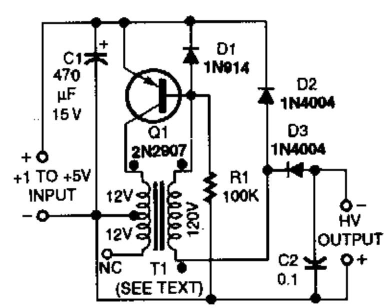
VOLTAGE TRIPLER FOR RADON DETECTOR IONIZATION CHAMBER



ELECTRONICS NOW *Fig. 75-2*

The voltage tripler charges the ionization chamber capacitor. It is powered from the 120-Vac line. Warning: Shock hazard exists.

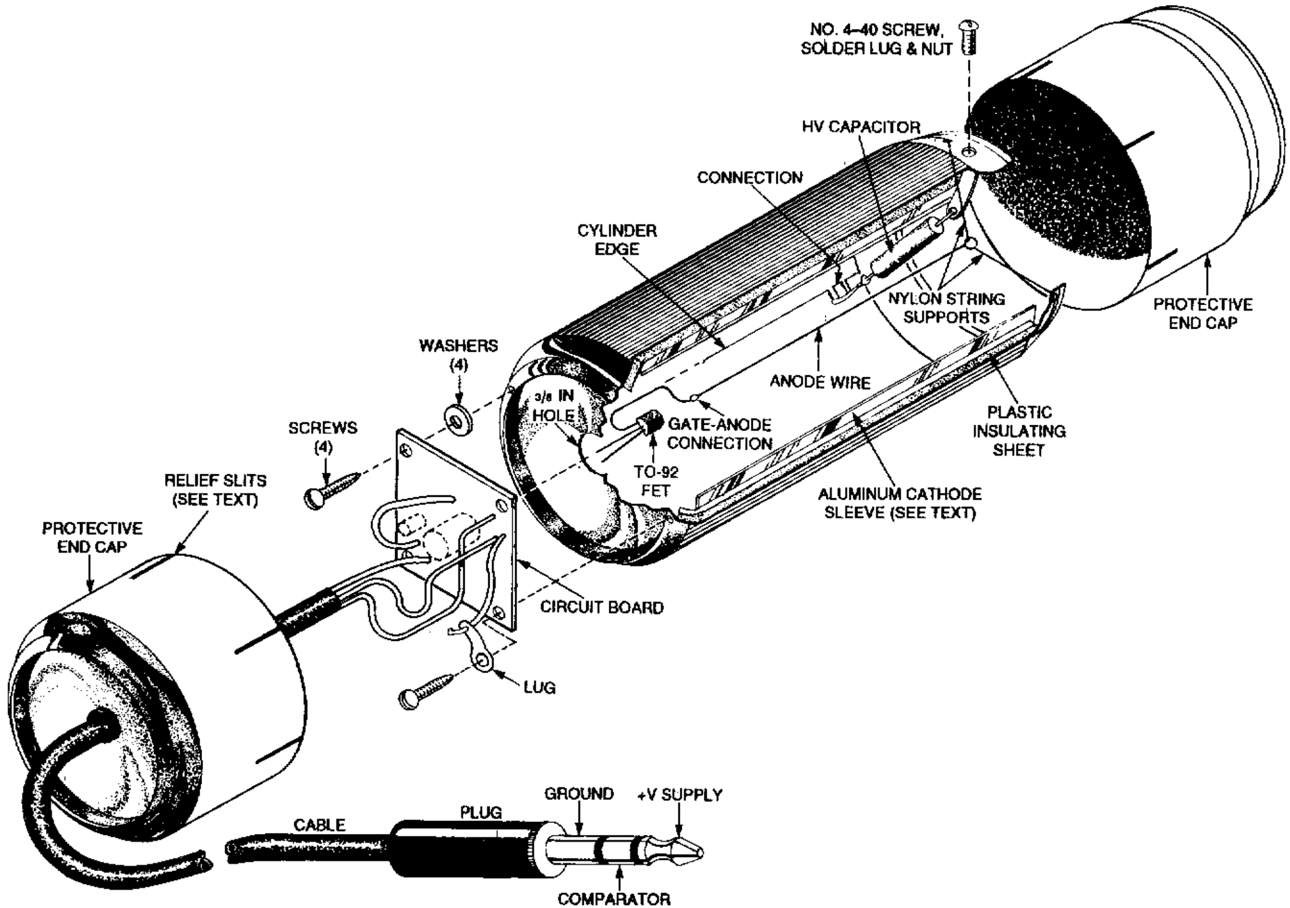
FLYBACK POWER SUPPLY FOR RADON MONITOR

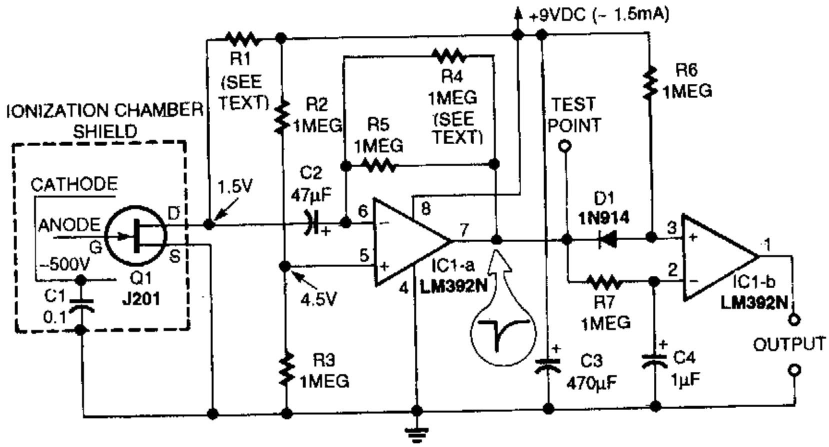


ELECTRONICS NOW *Fig. 75-3*

This blocking-oscillator flyback circuit is an alternative for charging the ionization chamber capacitor.

RADON MONITOR AMPLIFIER AND HEAD





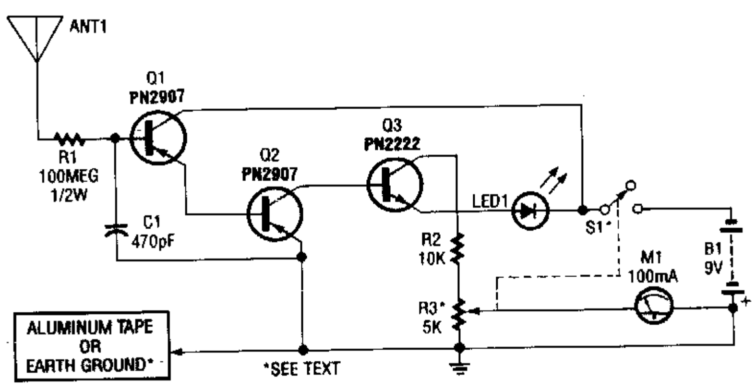
ELECTRONICS NOW

Fig. 75-4

A positively charged anode wire attracts electrons and a negatively charged cathode attracts positively charged ions. The recombination of electrons and ions causes a current that produces a voltage pulse. The cathode is maintained at –500 V by a charge on the 0.1-µF capacitor.

A beverage can forms the chamber, an aluminum can forms the cathode, and half cans form protective end covers. The amplifier circuit board is shown to the left of center.

ION DETECTOR



POPULAR ELECTRONICS

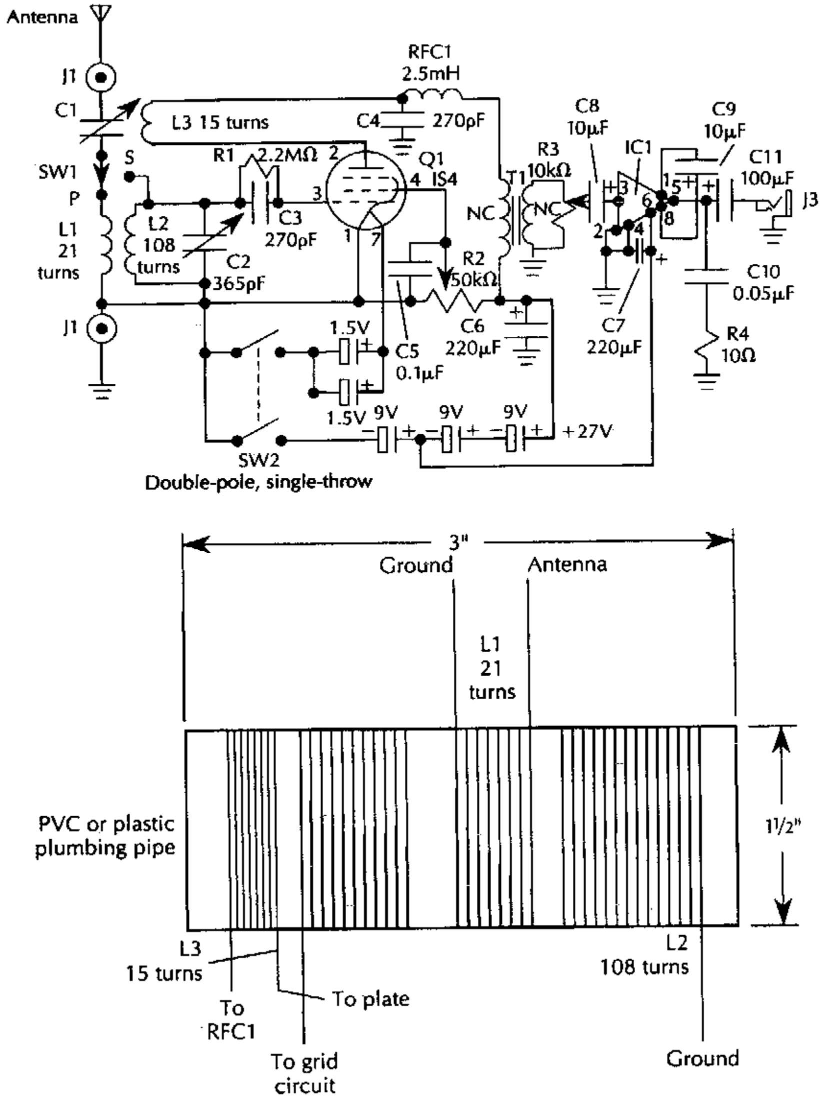
Fig. 75-5

ANT1 is a short whip antenna from a junked radio or other device. R_3 is adjusted to bring the meter on scale. This device should be grounded to operate properly. A length of aluminum or copper foil tape attached to the instrument case makes contact with the hand, and the body serves as a ground via hand contact with this tape.

76

Receiving Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

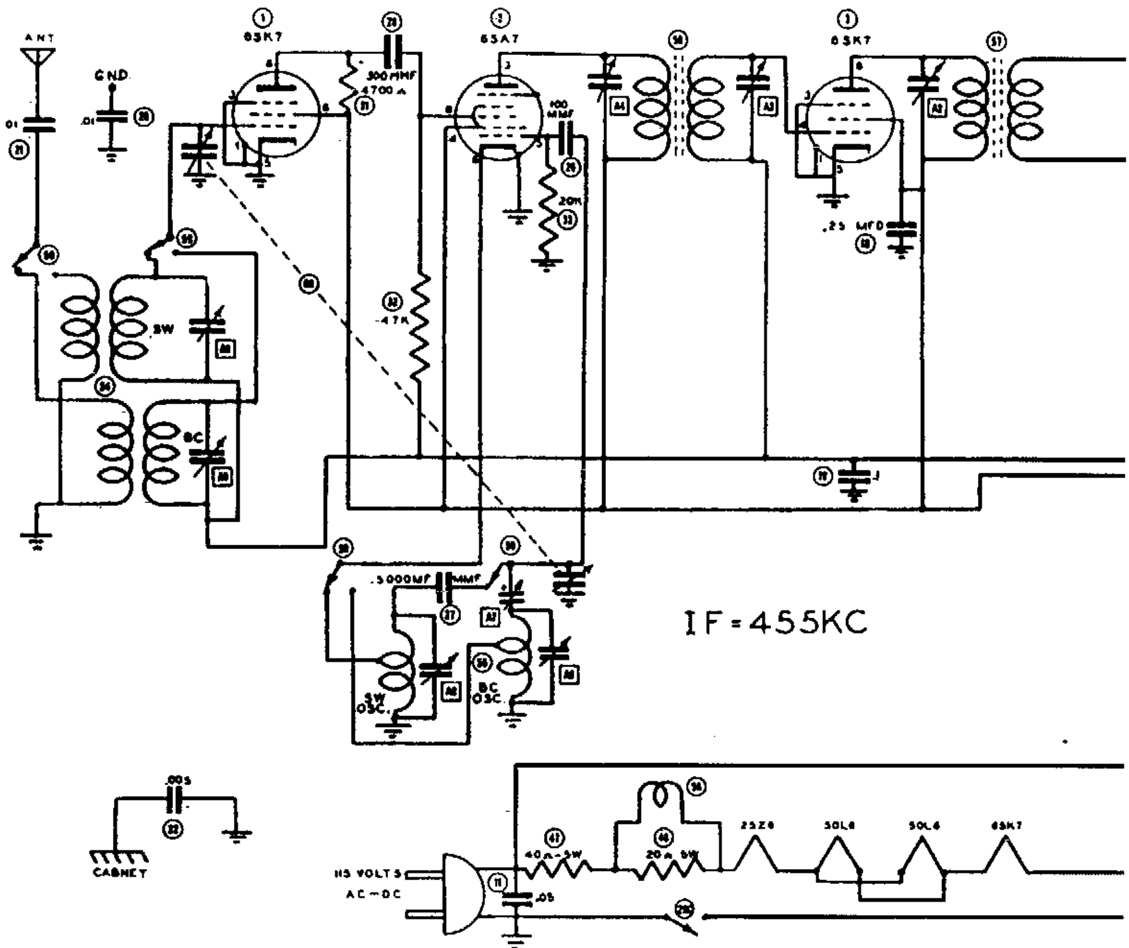


The primary winding of L1 has 21 turns of #24 or #26 enameled wire; L2 has 108 turns; and L3 has 15 turns of wire. All are wound on 1½-inch PVC pipe form.

McGRAW-HILL *Fig. 76-1*

A 1S4 regenerative detector feeds an LM386 audio IC (IC1). 1.5-V D cells and three 9-V batteries are used for a power supply.

ac/dc VACUUM-TUBE AM AND SHORTWAVE RECEIVER



POPULAR ELECTRONICS

This circuit was used in a World War II vintage AM/SW (6 to 18 MHz) receiver and shows typical circuits used in receivers at that time.

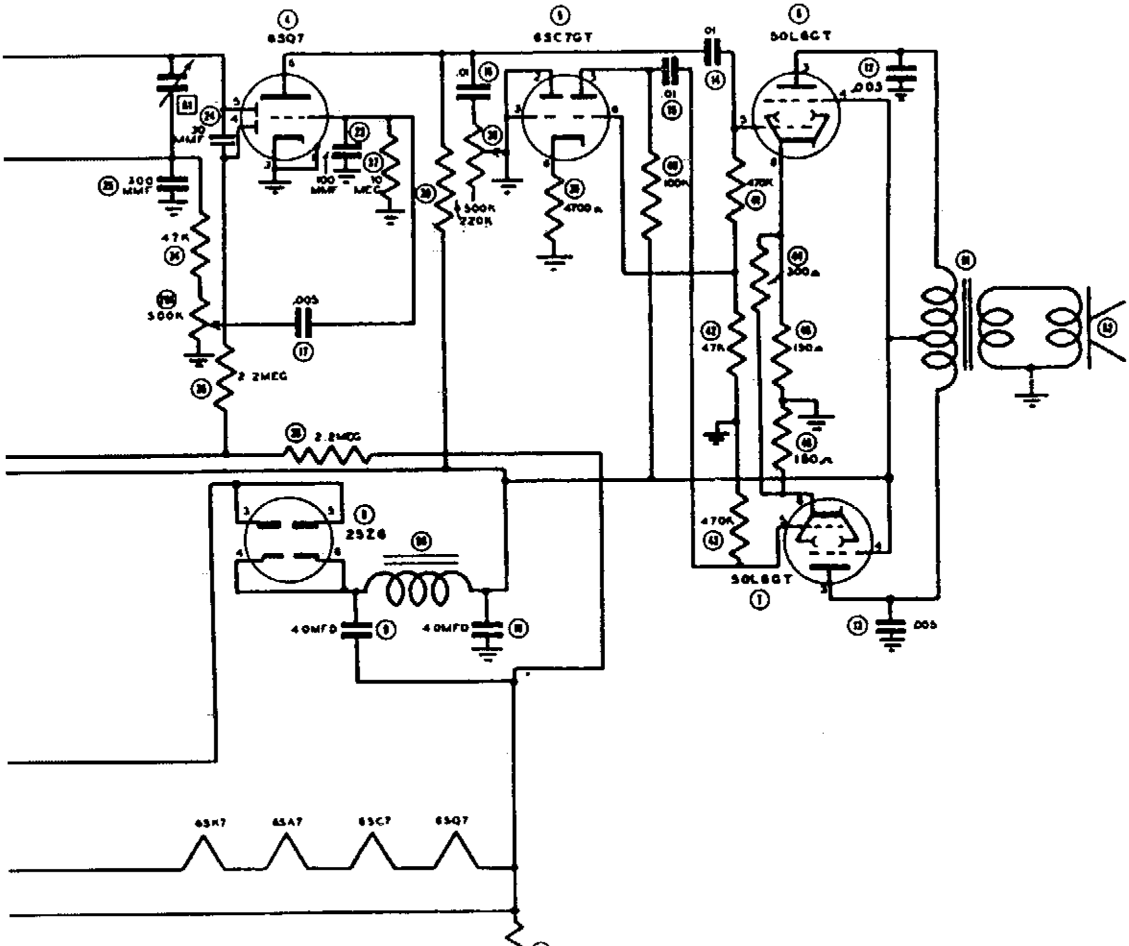

6SQ7
6SC7GT
50L6GT
25Z6
50L6GT
300 MMF
47K
500K
.005
2.2MEG
2.2MEG
500K
220K
4700
100K
470K
47K
470K
300
150
150
.005
.005
40MFD
40MFD
100 MMF
10 MEG
30 MMF
6SK7
6SA7
6SC7
6SQ7

Fig. 76-2

WWV RECEIVER

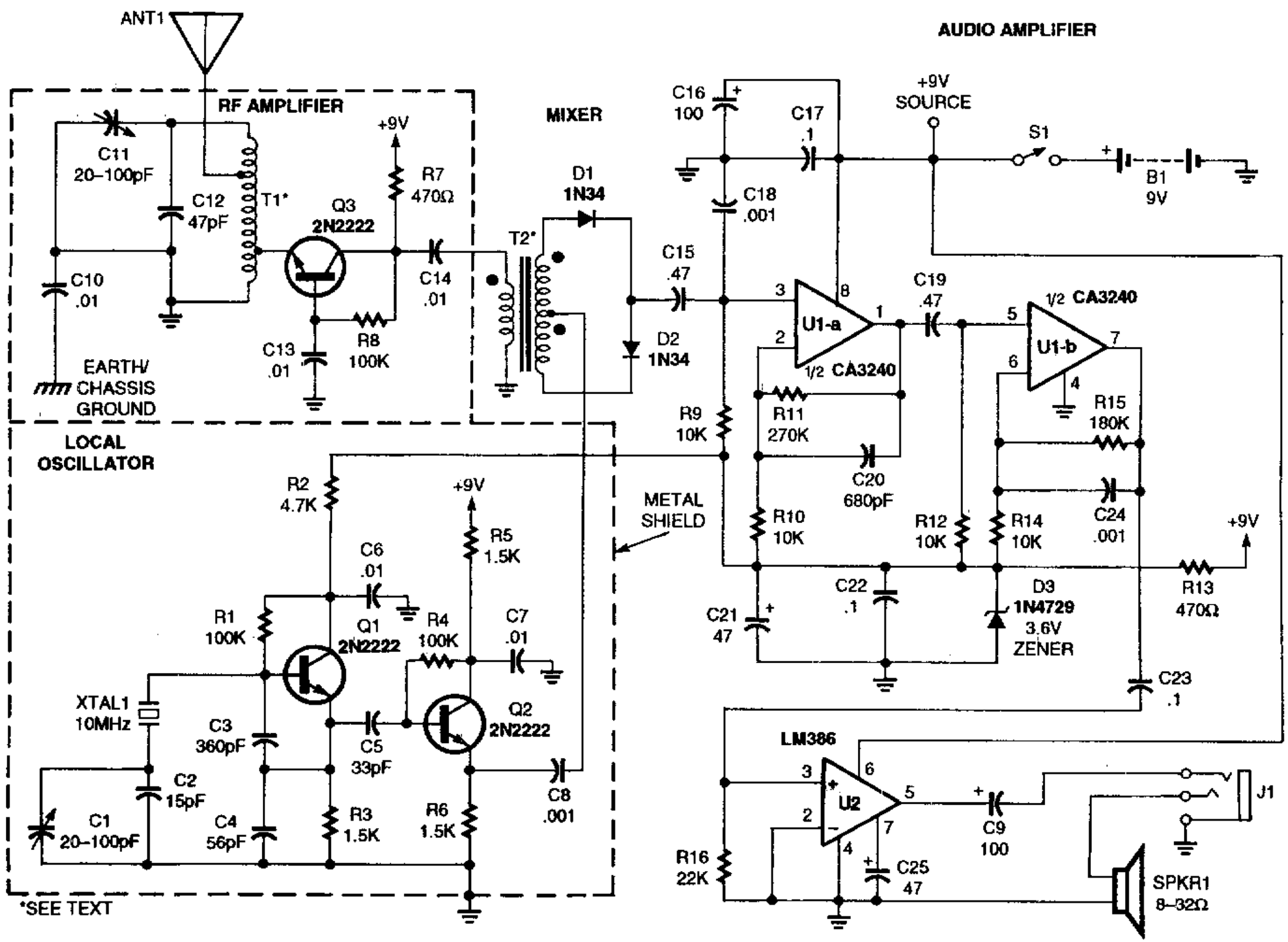
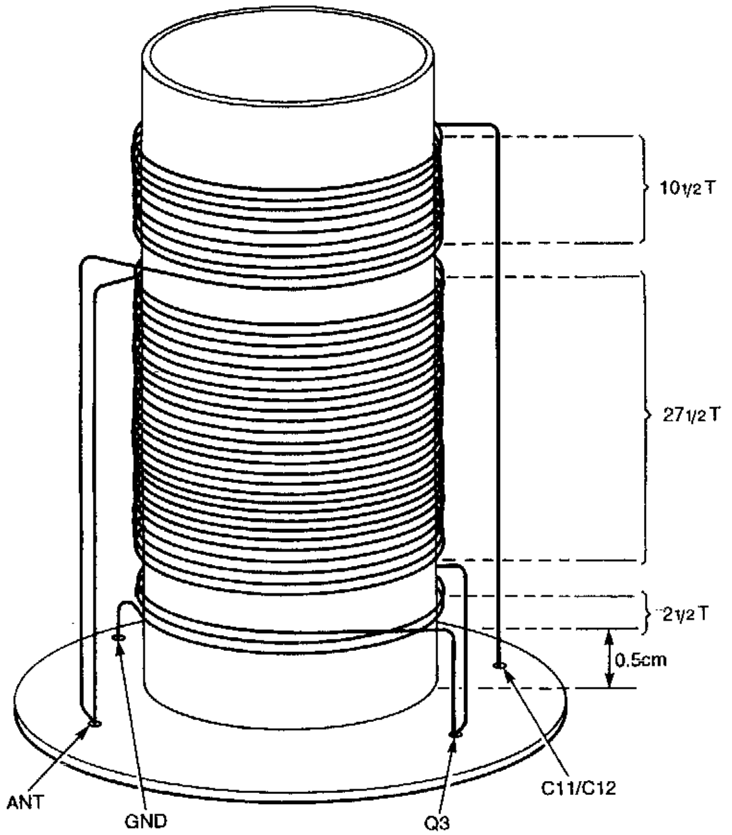
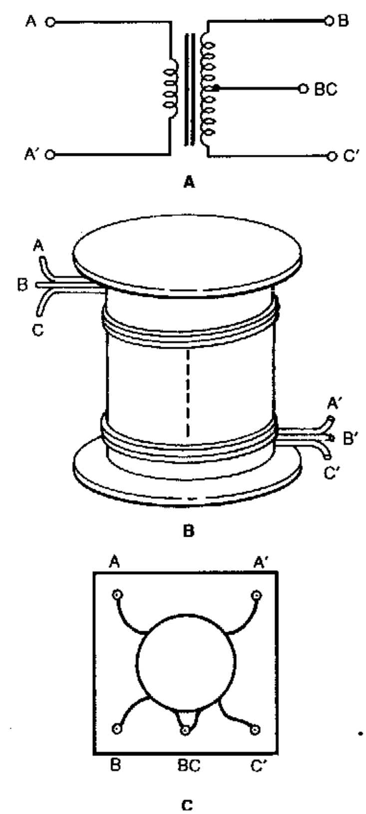


Fig. 76-3



Transformer T1 is a home-made unit comprised of 40 closely wound turns of #26 AWG enameled wire on a ¼-inch diameter air-core form, with taps at 2½ and 10½ turns from each end; after each tap, the winding continues in the same direction.



Construction details for T2 (the mixer transformer) are shown here. The diagram in A is a schematic representation of the unit once completed; B illustrates how the three lengths of wire are wound as a set on the bobbin; and C shows how the bobbin is connected to the pinned base.

RF amplifier Q3 feeds diode mixer D1–D2 and Q1–Q2 provide 10-MHz L.O. injection to D1 through T1 and T2. U1A, U1B and U2 are audio amplifiers. Details of T1 and T2 are shown.

SHORTWAVE RECEIVER

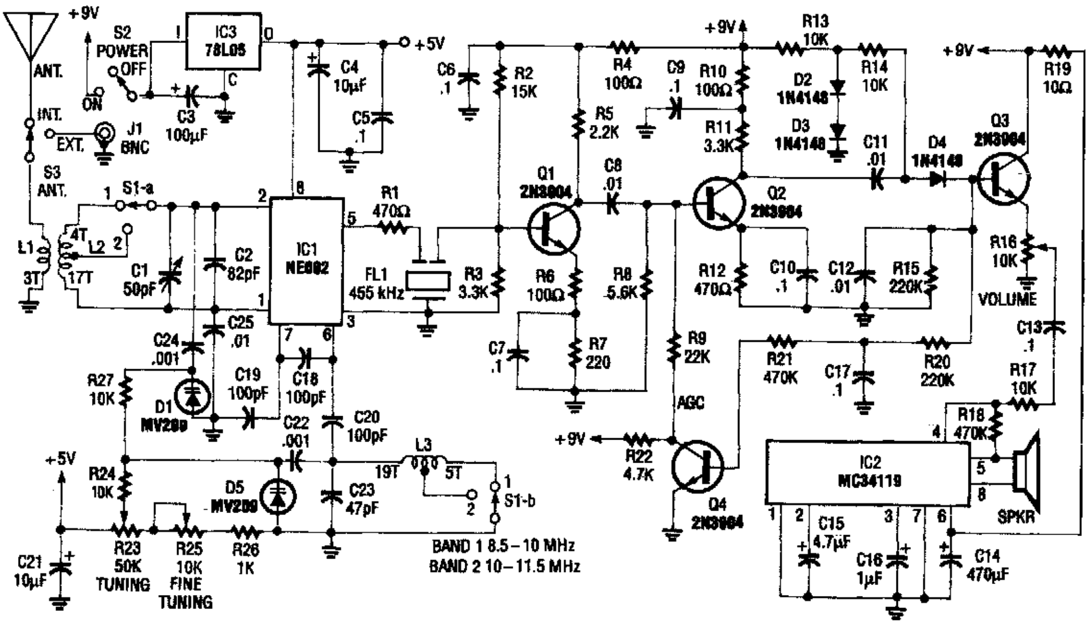


TABLE 2

Frequency (MHz)	C2 (pF)	C18, C19 (pF)	C23 (pF)	L1, L2 (Ant) (# of turns on T-37-2 core)	L3 (Osc)
5	100	120	68	5, 41	45
6	100	120	68	4, 30	34
7	82	100	47	4, 26	29
8	82	100	47	3, 22	24
10	82	100	47	3, 17	19
12	82	100	47	2, 15	17
14	68	82	33	2, 14	15
15	68	82	33	2, 13	14

TABLE 3

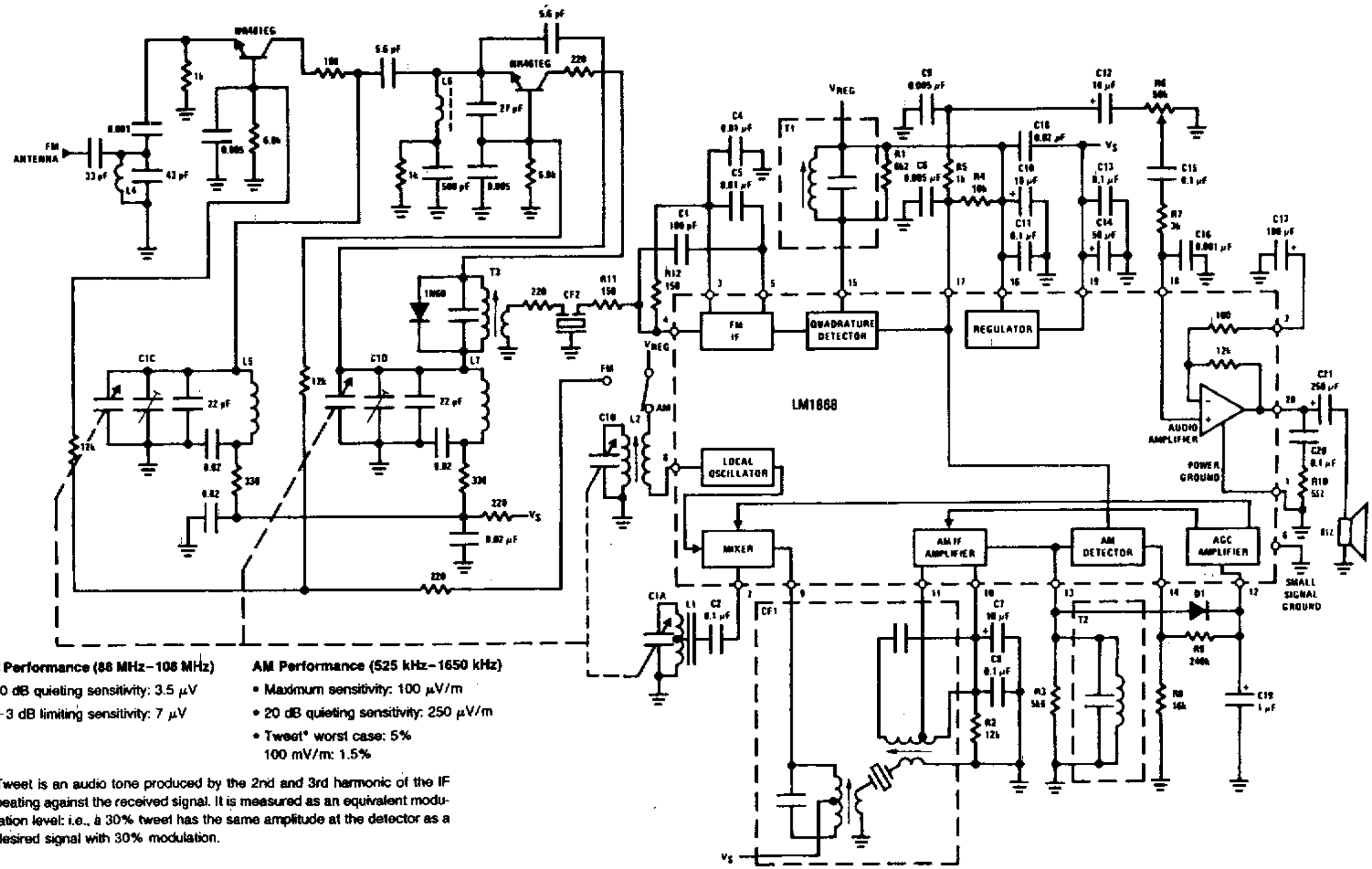
	NE602	MC34119
Pin 1	1.27 V	0V
Pin 2	1.27 V	4.15 V
Pin 3	0V	4.11 V
Pin 4	3.64 V	3.97 V
Pin 5	3.59 V	4.14 V
Pin 6	4.99 V	-9.09 V
Pin 7	4.33 V	0V
Pin 8	5.05 V	4.20 V

	Q1	Q2	Q3	Q4
Emitter	0.95 V	0.80 V	0.27 V	0V
Base	1.61 V	1.45 V	0.82 V	0.58 V
Collector	2.56 V	3.30 V	9.17 V	7.41 V

RADIO-ELECTRONICS *Fig. 76-4*

This receiver covers 8.5 to 11.5 MHz in two bands and has a sensitivity of under 1 µV. An NE602 mixer feeds a 455-kHz IF amplifier (Q1 and Q2), detector D4, and audio amplifier IC2. Q4 serves as an AGC amplifier coil data is given in the table. The LO is varactor tuned.

AM/FM RECEIVER CIRCUIT



FM Performance (88 MHz–108 MHz)

- 30 dB quieting sensitivity: 3.5 μV
- −3 dB limiting sensitivity: 7 μV

AM Performance (525 kHz–1650 kHz)

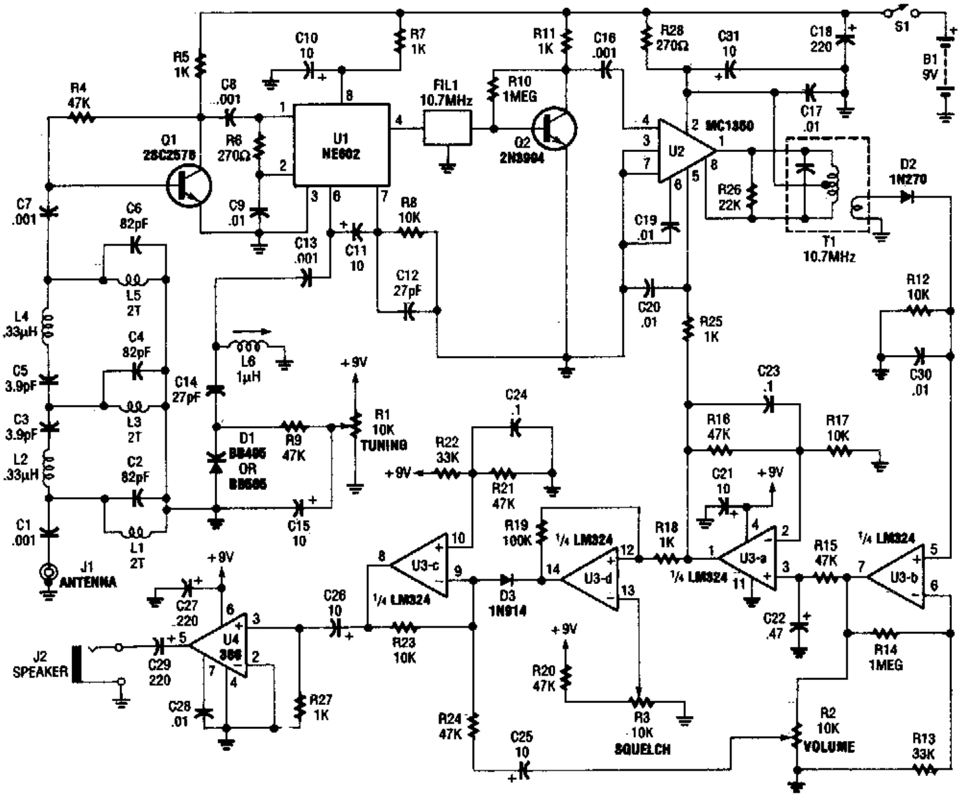
- Maximum sensitivity: 100 μV/m
- 20 dB quieting sensitivity: 250 μV/m
- Tweet* worst case: 5%
  100 mV/m: 1.5%

*Tweet is an audio tone produced by the 2nd and 3rd harmonic of the IF beating against the received signal. It is measured as an equivalent modulation level: i.e., a 30% tweet has the same amplitude at the detector as a desired signal with 30% modulation.

NATIONAL SEMICONDUCTOR *Fig. 76-5*

This circuit shows the LM1868 as a complete AM radio and FM IF section. An external FM front end is used for the 88- to 108-MHz band. Audio output is 0.5 W and either 9-V battery or line operated supply can be used.

118- TO 136-MHz AIRCRAFT RECEIVER

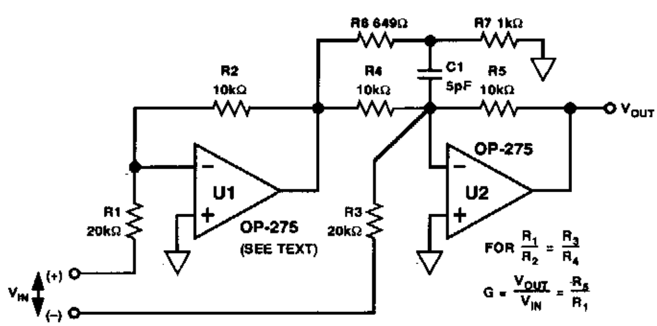


RADIO CRAFT

Fig. 76-6

This receiver covers the 118- to 136-MHz AM aviation band. It has a 10.7-MHz IF amplifier. L1, L3, and L5 are 1½ turns of #24 wire. F1L1 is a 10.7-MHz ceramic filter. IF bandwidth will be about 250 kHz.

DUAL-INVERTER LINE RECEIVER

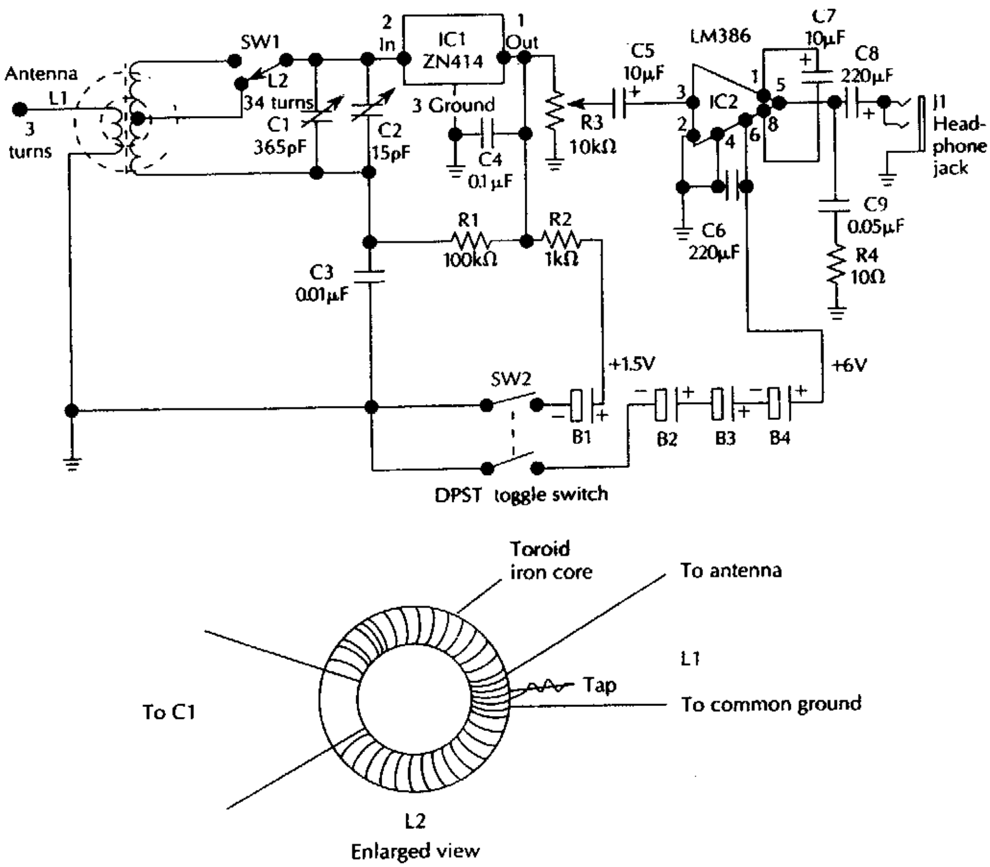


ANALOG DEVICES

Fig. 76-7

This circuit is for audio applications.

TOROIDAL CORE TRF SHORTWAVE RECEIVER



L1—3 turns of #24 enameled magnet wire wound over center of L2.
L2—34 turns of #24 enameled wire—coils wound on toroid iron core form T-50-2—tapped at the 17th turn. Form has only 1/2-inch diameter.

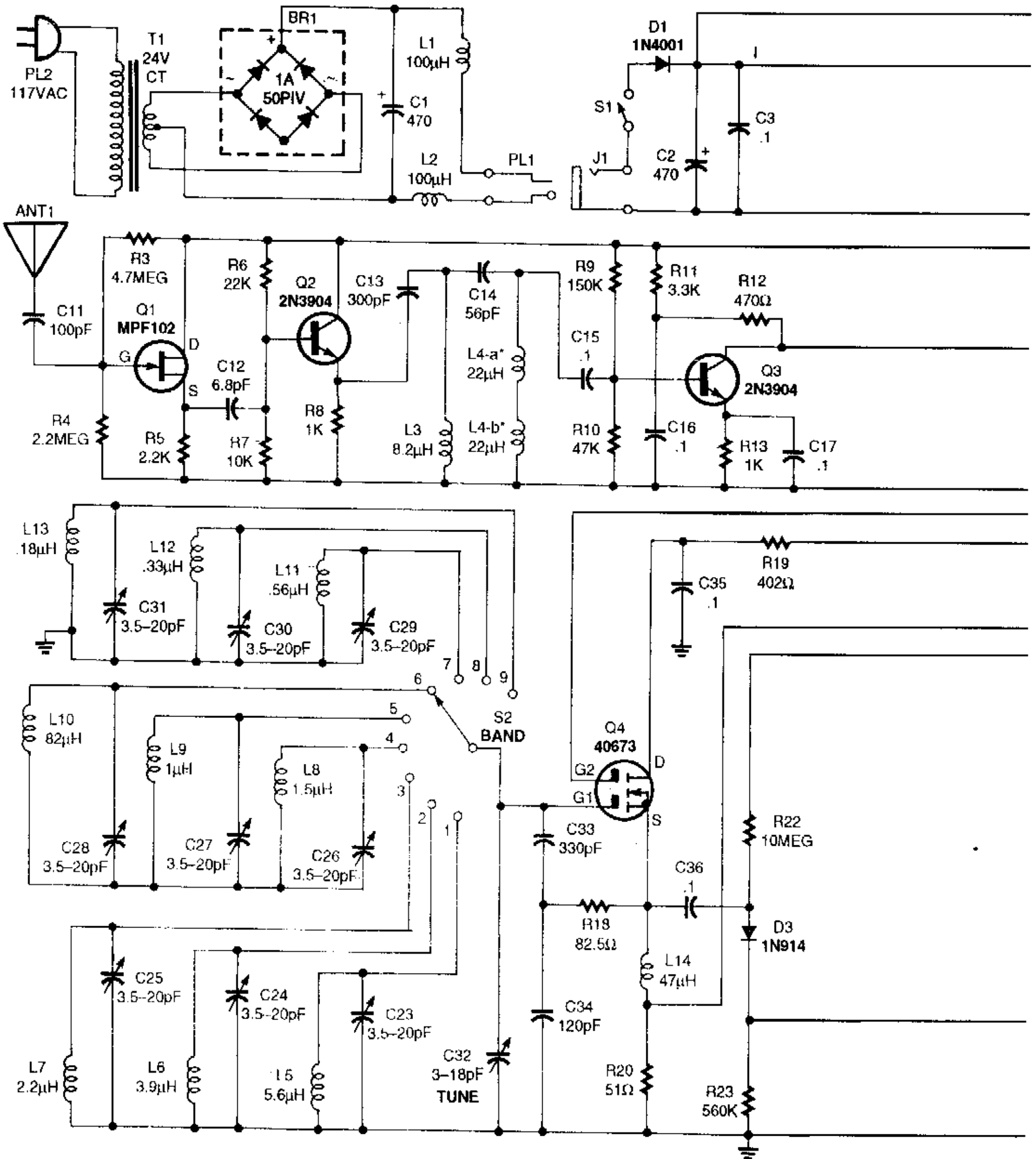
Wind 34 turns, with a tap at the 17th turn, for L2 on the small ½-inch diameter iron core form.

McGRAW-HILL

Fig. 76-8

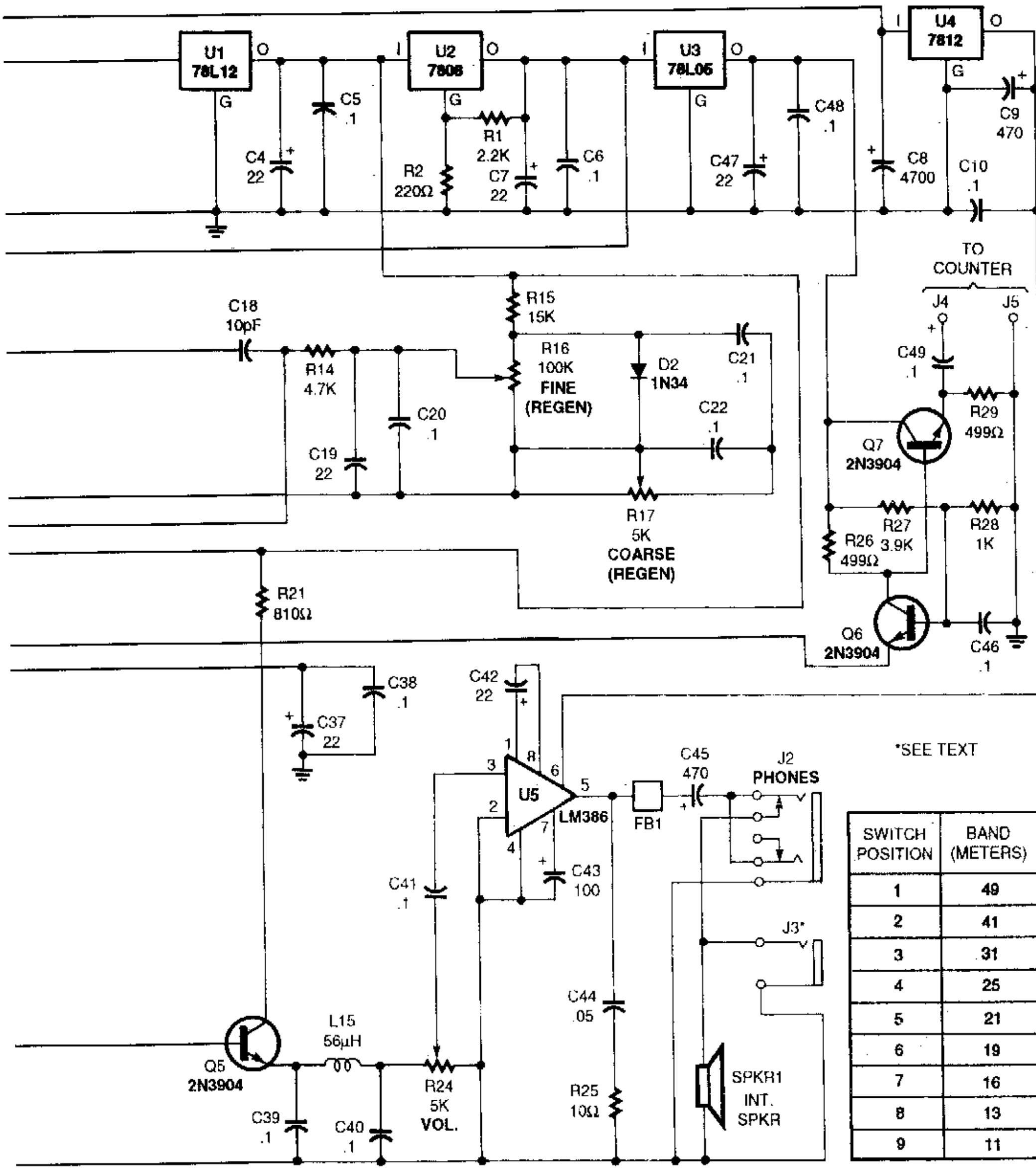
A ZN414 IC feeds an LM386 audio amplifier in this TRF circuit. SW1 is a band-switch. Coverage is up to 18 MHz.

NINE-BAND SHORTWAVE RECEIVER



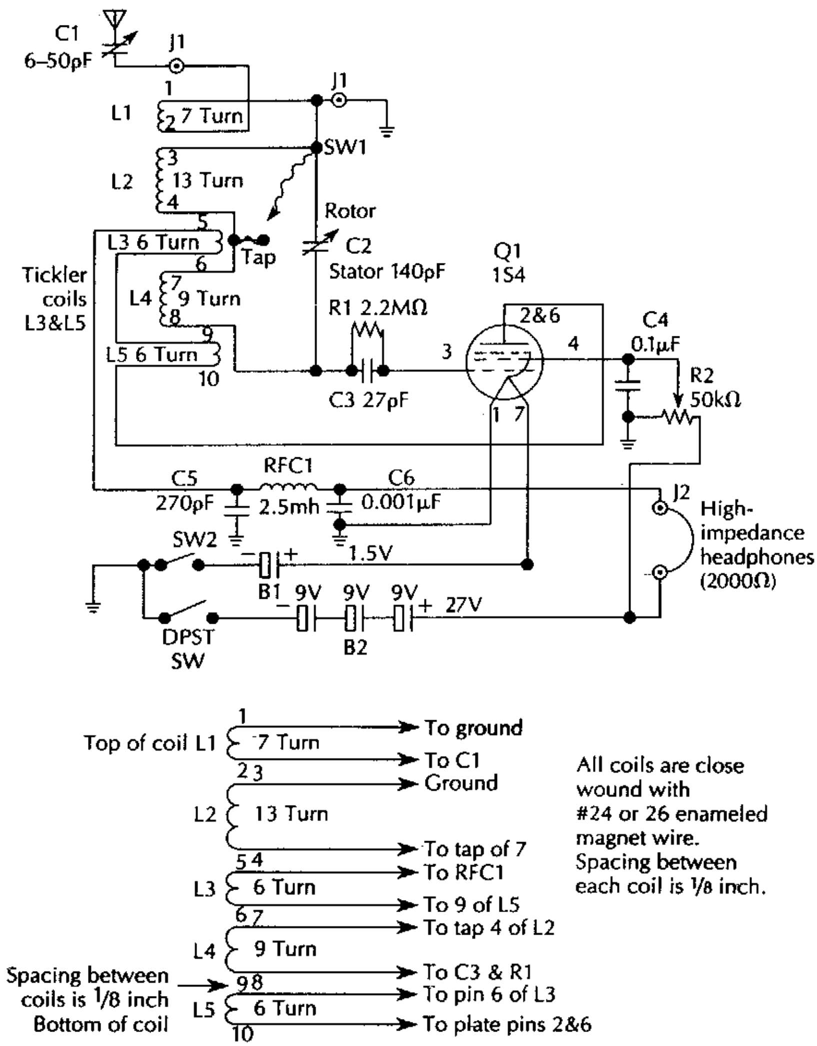
POPULAR ELECTRONICS

Dual-gate MOSFET Q4 is used as a regenerative amplifier in this circuit. An active antenna feeds the signal to Q4, and a short whip antenna is adequate. Detector Q5 feeds volume control R24, and audio amplifier U5, an LM386. The frequency range is 49 to 11 meters in nine bands (6 to 27 MHz).



SWITCH POSITION	BAND (METERS)
1	49
2	41
3	31
4	25
5	21
6	19
7	16
8	13
9	11

ONE-TUBE REGENERATIVE SW RECEIVER

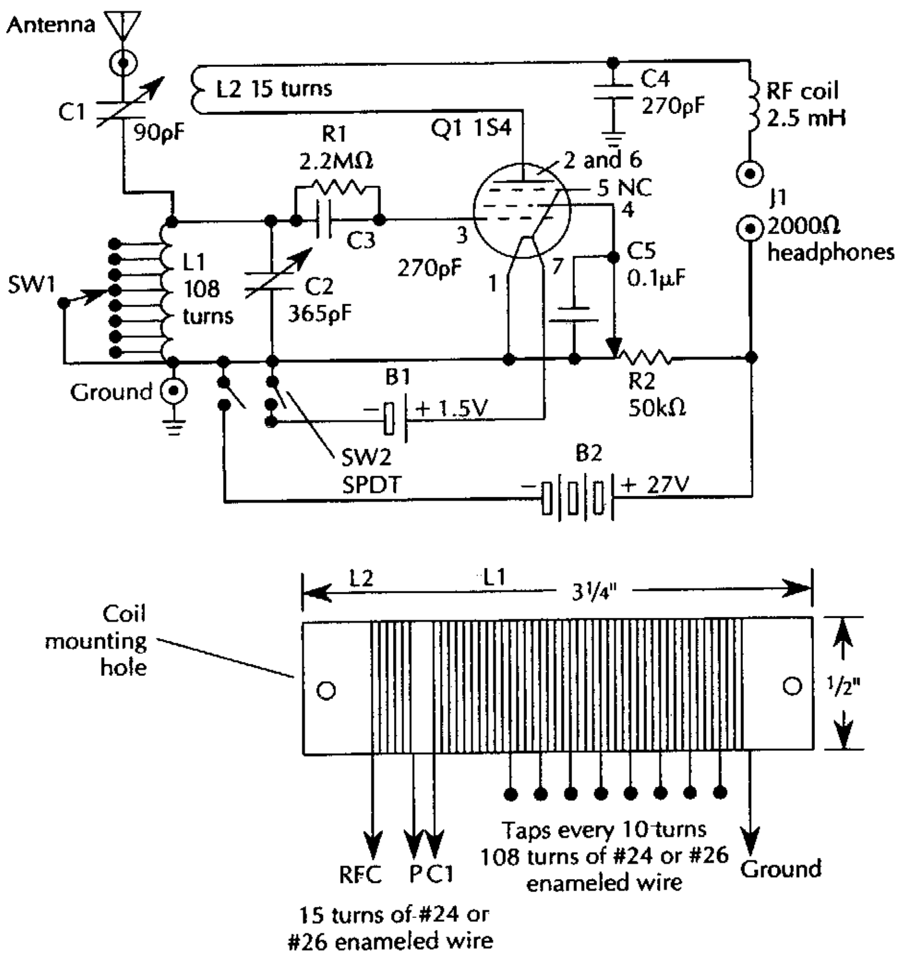


How to wind L1, L2, and L3 with taps. This receiver tunes in the 40- and 80-meter range of frequencies.

McGRAW-HILL ***Fig. 76-10***

A 154 tube is used in a regenerative detector circuit. Details for coils are shown and frequency range can be shifted within 1.5 to 20 MHz by proportionally adjusting the number of turns on coils.

ONE-TUBE REGENERATIVE AM RECEIVER



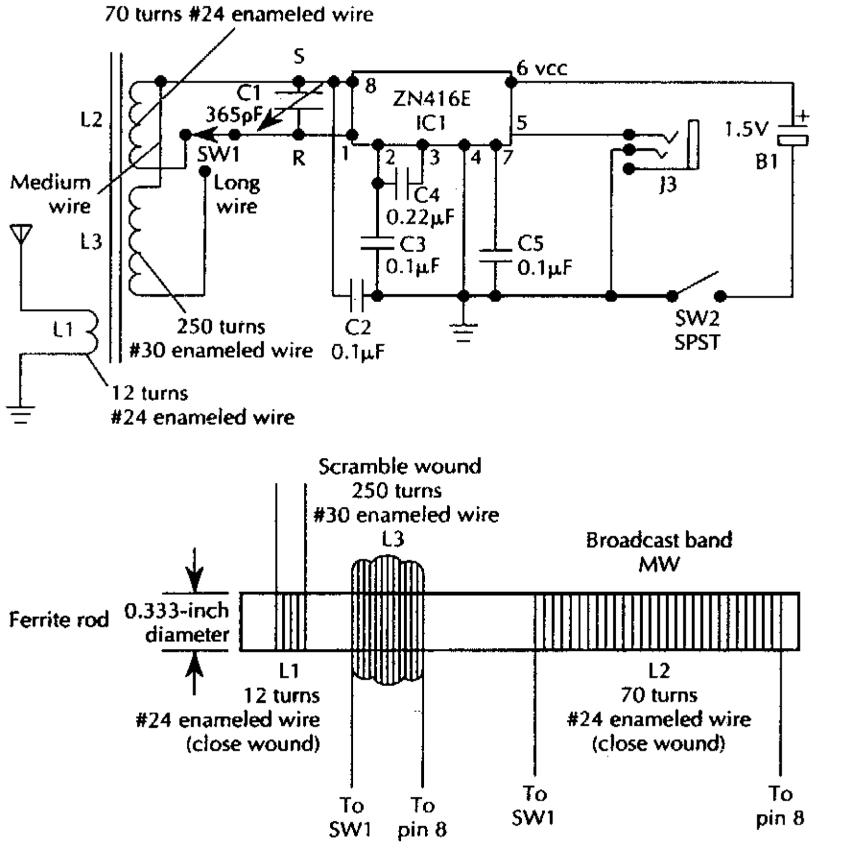
Wind both coils on PVC pipe using either #24 or #26 enameled wire. For coil L1, wind 108 turns on the pipe, and tap every 10 turns.

McGRAW-HILL

Fig. 76-11

Suitable for AM reception and as a simple radio project, this circuit uses a single tube as a regenerative detector.

TWO-BAND RADIO

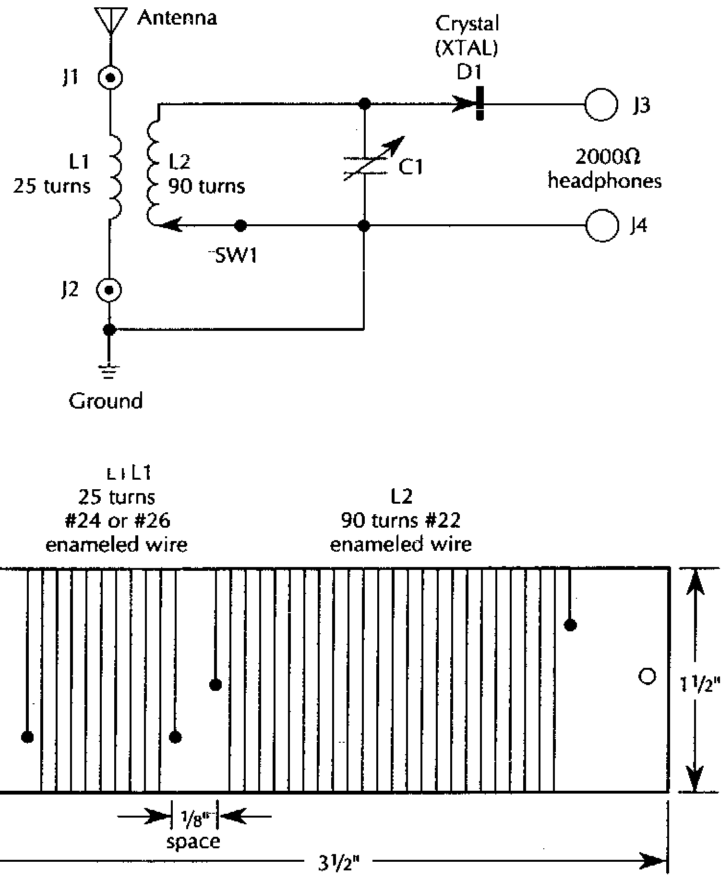


All three coil windings are wound on one long ferrite form. Two different coils are switched into the circuit, covering the longwave (lw) and medium-wave (mw) broadcast band.

MCGRAW-HILL *Fig. 76-12*

This TRF receiver covers the AM broadcast band and longwave bands (used in Europe and Asia for broadcasting). A loop antenna is used for reception and an external antenna can be connected. Frequency coverage is 150 to 1600 kHz.

SIMPLE CRYSTAL RADIO

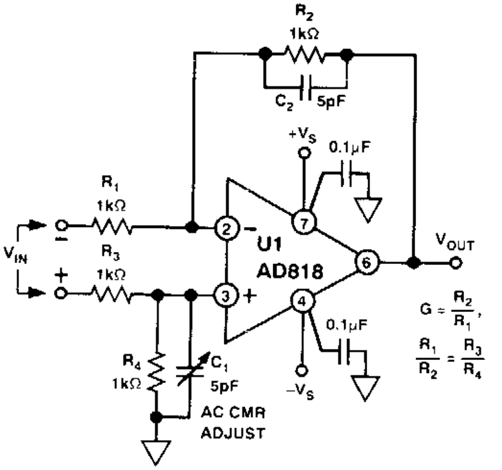


Wind 90 turns or 2½ inches of #22 enameled wire for L2, and 25 turns of #24 or #26 enameled wire for L1 on a 1½-inch PVC plastic pipe form.

McGRAW-HILL *Fig. 76-13*

An IN34A (D1) is used as a detector in this crystal radio. A good outdoor antenna should be used.

VIDEO LINE RECEIVER

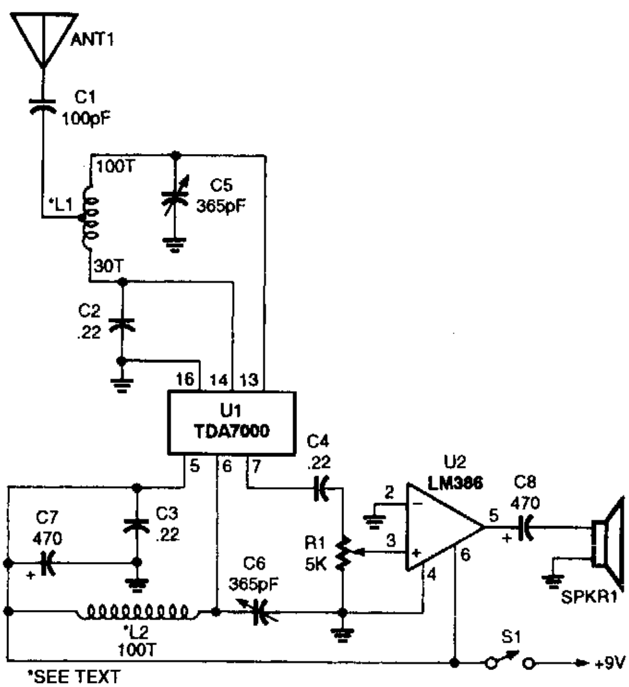


This circuit can achieve 46-dB common-mode rejection if R1, R2, R3, and R4 are matched to 1%. C1 is adjusted for best CMR above 1 MHz.

ANALOG DEVICES

Fig. 76-14

TWO-CHIP AM RECEIVER

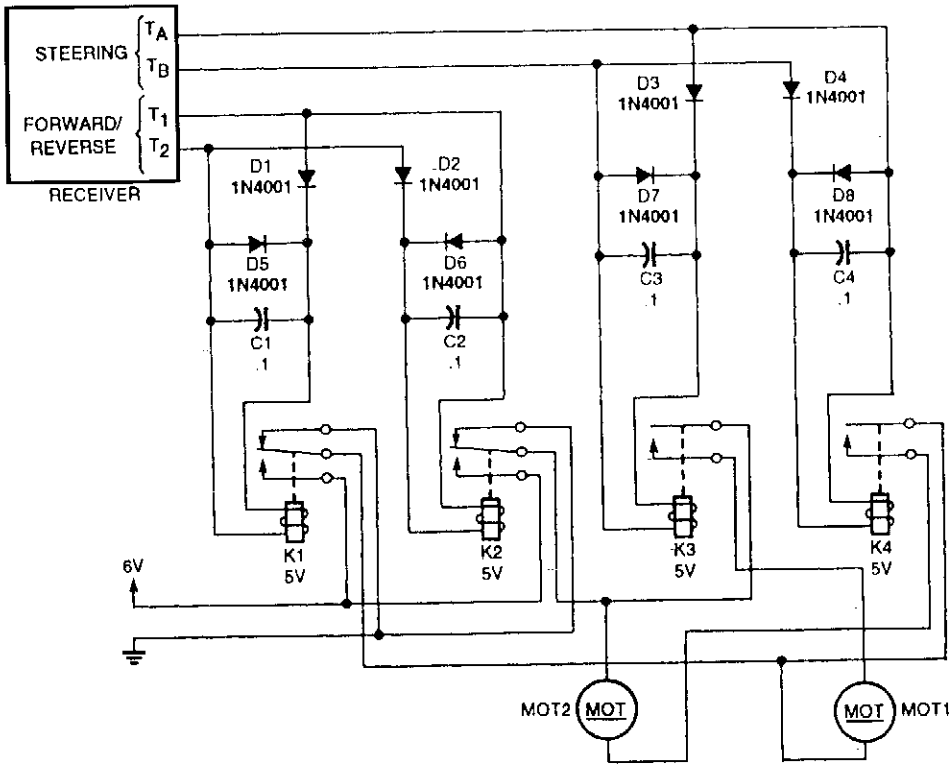


POPULAR ELECTRONICS

Fig. 76-15

This receiver is comprised of a TDA7000 single-chip FM receiver (U1), an LM386 low-voltage audio-power amplifier (U2), a pair of hand-wound coils (L1 and L2), and a few additional components. L1 and L2 are 100 turns of #28 wire on toroidal cores (about 240 μH each). L1 is tapped at 30 turns.

RELAY INTERFACE TO RC RECEIVERS

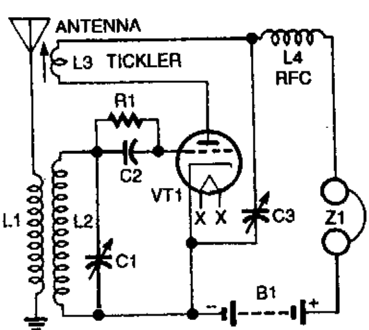


POPULAR ELECTRONICS

Fig. 76-16

You can add relays to some inexpensive RC receivers to operate your own chassis.

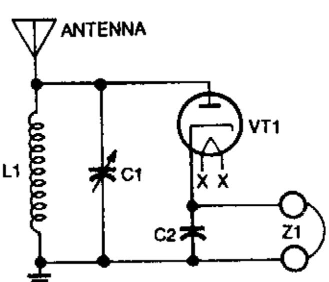
BASIC REGENERATIVE RECEIVER



ELECTRONICS NOW

Fig. 76-17

SIMPLE RADIO RECEIVER

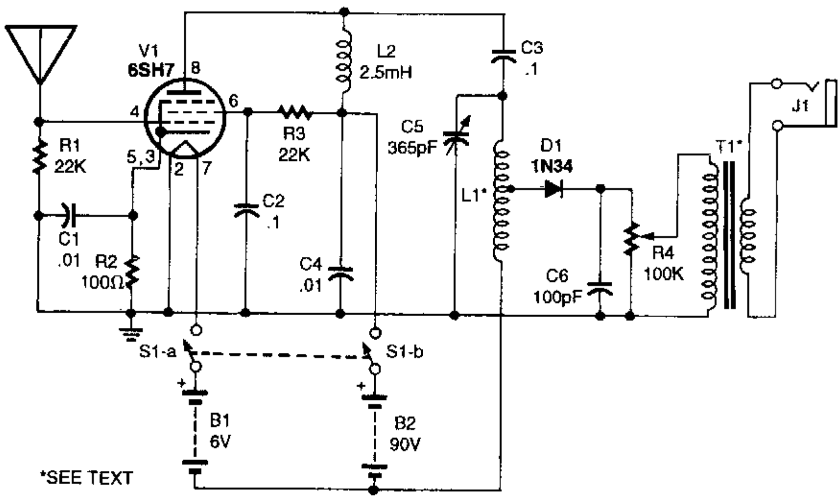


ELECTRONICS NOW

Fig. 76-18

Vacuum-tube detector receiver.

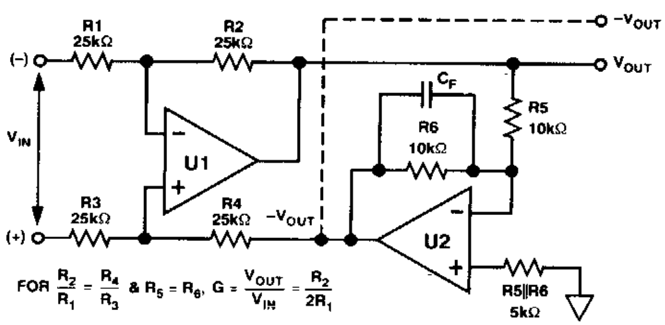
ONE-TUBE AM RECEIVER



POPULAR ELECTRONICS *Fig. 76-19*

This radio uses an untuned RF stage to boost the signal voltage up to the linear portion of the crystal diode's characteristic curve. The circuit's distortion and wide bandpass and a good-quality transformer make for a great-sounding AM radio. L1 is a winding of #22 enamelled wire 2" long on a 2" diameter plastic pipe. T1 is a tube-type radio output transformer, rated at 2000 Ω to the speaker voice coil.

BALANCED LINE RECEIVER

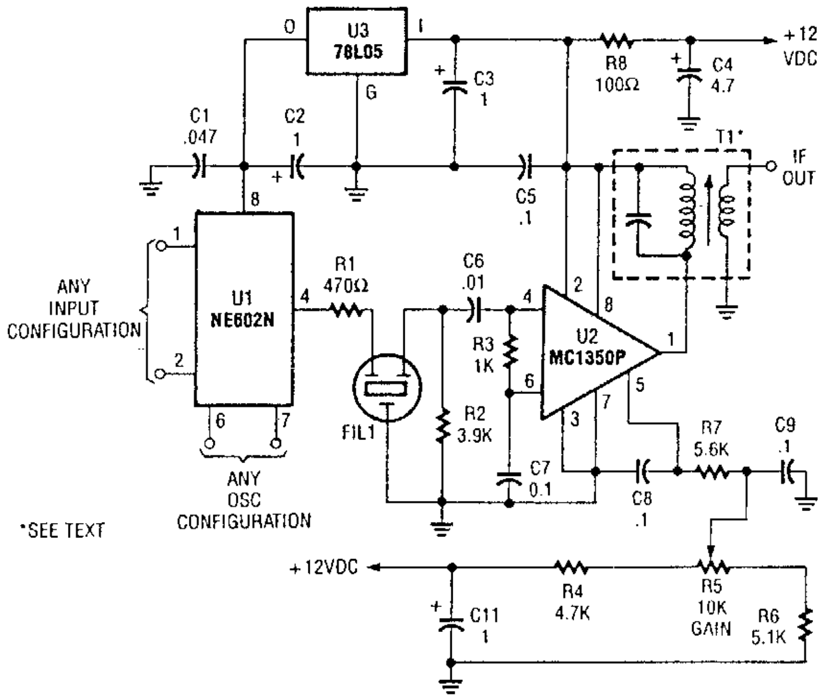


FOR $\frac{R_2}{R_1} = \frac{R_4}{R_3}$ & $R_5 = R_6$, $G = \frac{V_{OUT}}{V_{IN}} = \frac{R_2}{2R_1}$

ANALOG DEVICES *Fig. 76-20*

Unity-gain inverter U2 drives R4 (usually grounded at $-V_{out}$, equalizing currents in ±input legs, and provides a choice of balanced p-p output with a gain of R_2/R_1).

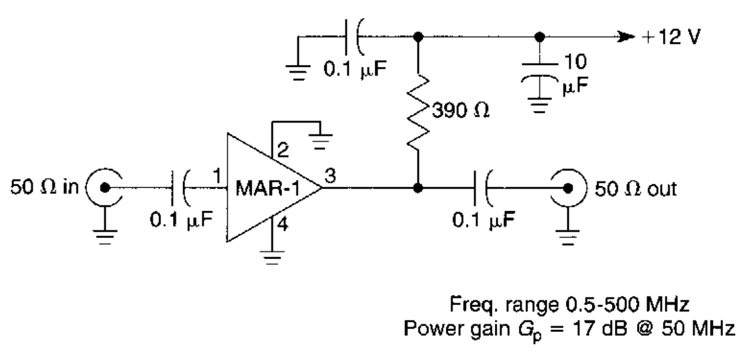
SUPERHET FRONT END



RADIO CRAFT *Fig. 76-21*

This superhet receiver front end is simple and uses an NE602 followed by an MC1350 IF amplifier.

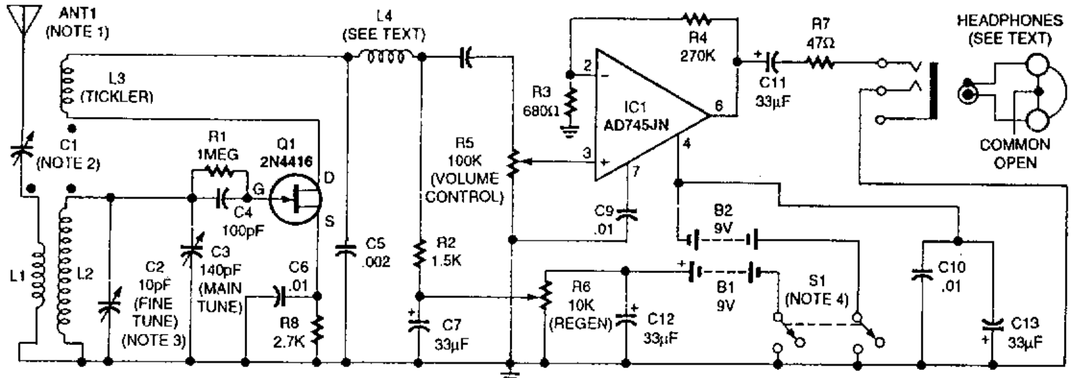
RECEIVER PREAMP



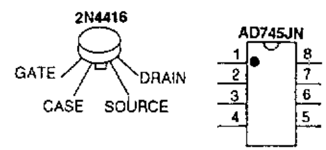
WILLIAM SHEETS *Fig. 76-22*

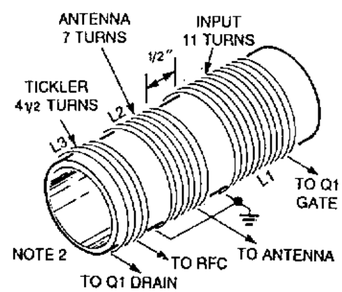
Suitable for HF and VHF receivers, this preamplifier can be mounted on the back of the receiver for a boost in gain. Useful gain is about 17 dB at 50 MHz.

REGENERATIVE RECEIVER FOR 6 TO 17 MHz



NOTES:
1. ANTENNA: 100FT COPPER WIRE
2. ADJUST C1 FOR ANTENNA USED
3. C2 OPTIONAL (SEE TEXT)
4. S1- DUAL ON OFF SWITCH.
5. FOR COIL-WINDING DIRECTIONS SEE FIG. 9-a





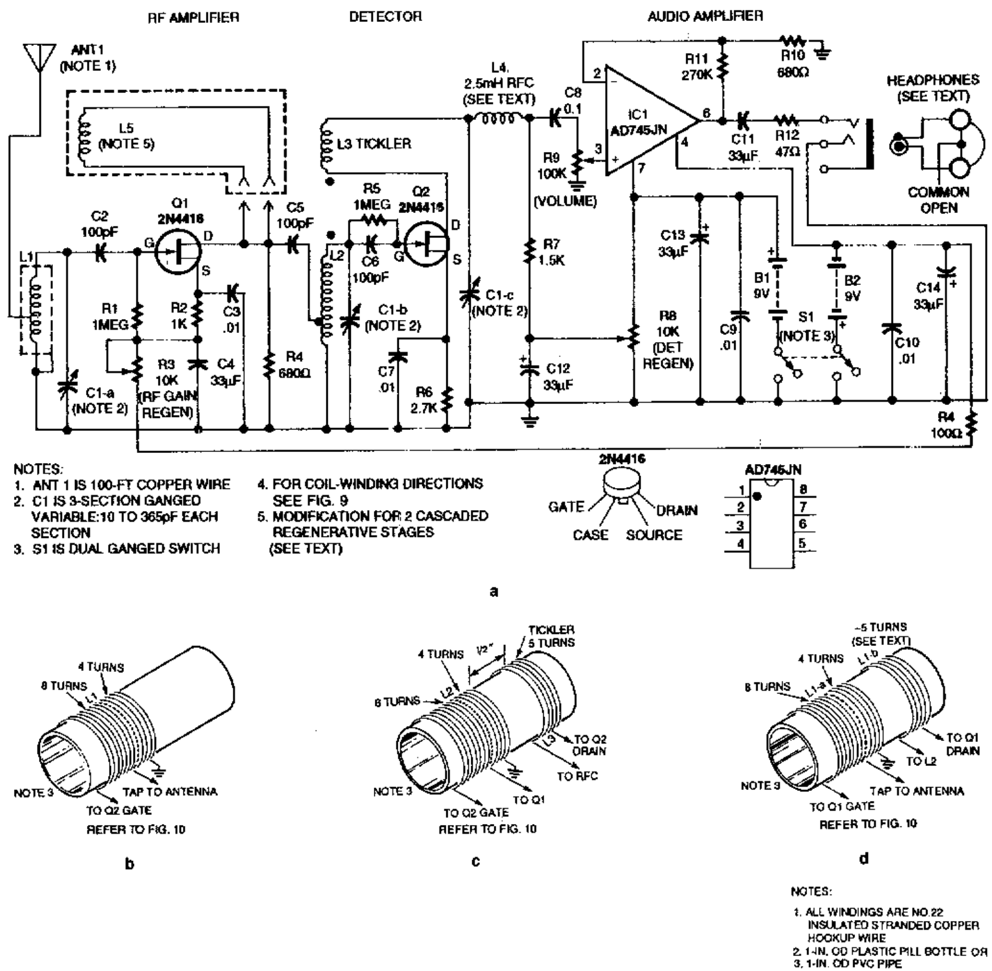
NOTES:
1. ALL WINDINGS ARE NO.22 INSULATED STRANDED COPPER HOOKUP WIRE
2. 1 1/2-IN. OD PVC PIPE

ELECTRONICS NOW!

Fig. 76-23

The headphones are 32-Ω stereo types. The common lead is left floating so that the two sides are in series, giving 64 Ω.

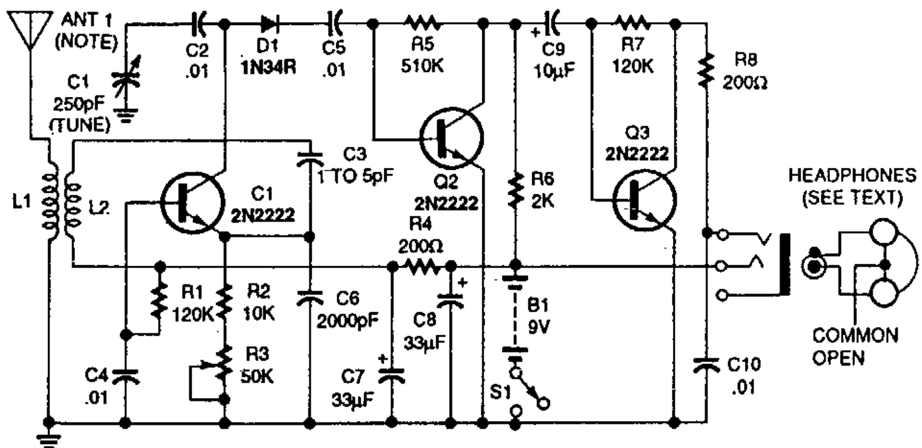
TWO-STAGE TRF REGENERATIVE RECEIVER



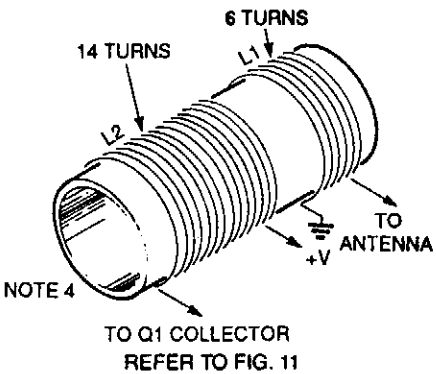
ELECTRONICS NOW *Fig. 76-24*

This regenerative receiver uses a tuned RF stage to improve performance. The coil in Fig. 76-24D is for the purpose of adding a second regenerative stage (RF amp). This coil is L5 in the schematic.

ECONOMY SHORTWAVE RECEIVER



1. ANT 1 IS 100-FT COPPER WIRE.
NOTES: 2 FOR COIL WINDING
DIRECTIONS SEE FIG. 9



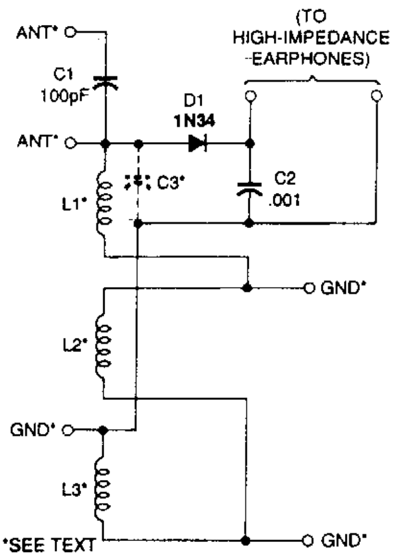
NOTES:

1. ALL WINDINGS ARE NO.22 INSULATED STRANDED COPPER HOOKUP WIRE
2. 1-IN. OD PLASTIC PILL BOTTLE
3. 1-IN. OD PVC PIPE OR NOTE 3

ELECTRONICS NOW *Fig. 76-25*

Using three transistors, this receiver covers the range of 6 to 17 MHz. Coils can be altered to change the range to a lower or higher frequency.

VARIOMETER-TUNED RADIO



POPULAR ELECTRONICS *Fig. 76-26*

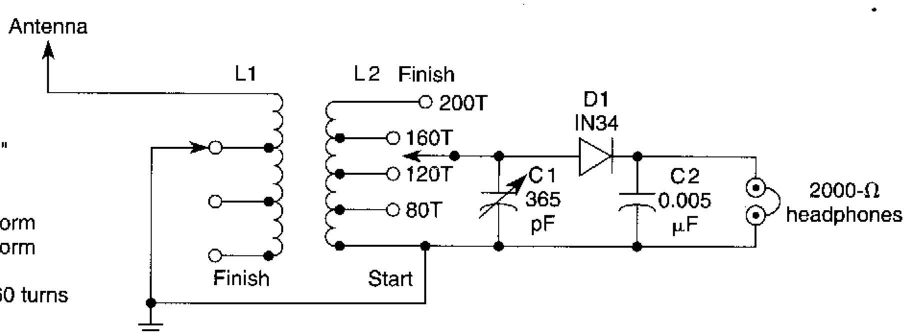
The two fixed coils of the variometer, L1 and L3, are wound on an 8½-inch-long piece of 1-inch-diameter plastic pipe (its outer diameter is about 1¼ inches). Each coil is 2¼ inches long. The number of turns is not critical, but 86 tightly wound turns of #22 enameled wire were used. When winding the coils, make sure you start at a point that will allow them to be placed 2 inches apart on the pipe. Drill holes in the pipe and run the leads of the coils out the end of the pipe that is closest to each.

The movable coil, L2, is wound on a piece of 1½-inch plastic pipe (its outer diameter is about 1⅞ inches). The winding is 2 inches long. Like L1 and L3, the actual number of windings of this coil are not critical, as long as the winding is approximately the right length.

OLD-FASHIONED CRYSTAL RADIO

Coil: Wound on 10" long 2" diameter form

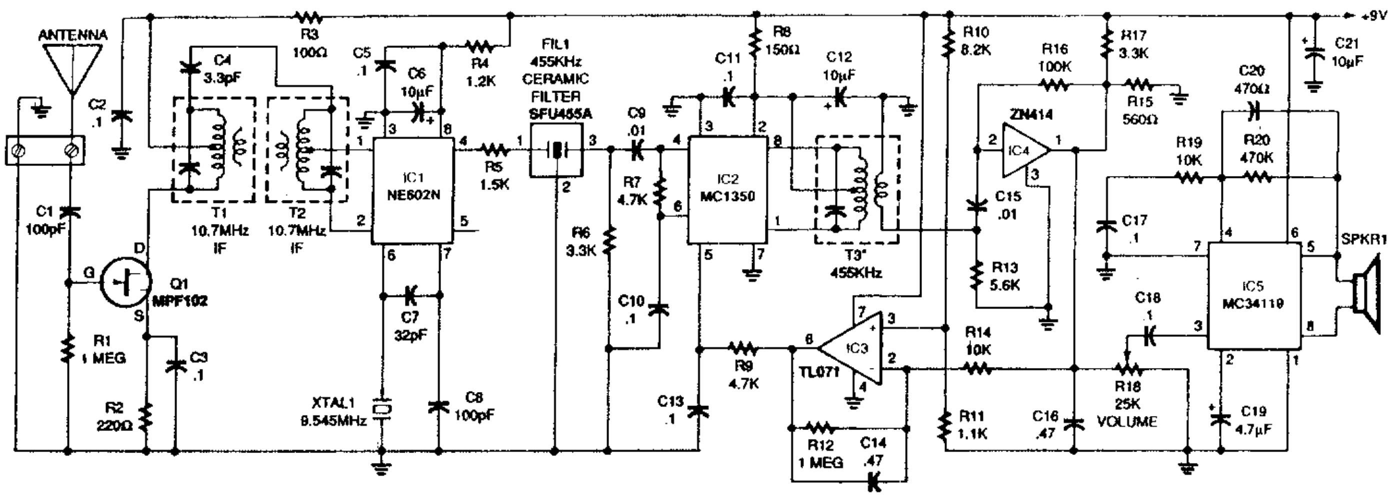
L1: 200T #24 on 2" dia. form
L2: 200T #24 on 2" dia. form spaced 1/4" from L1 tapped at 80, 120, 160 turns



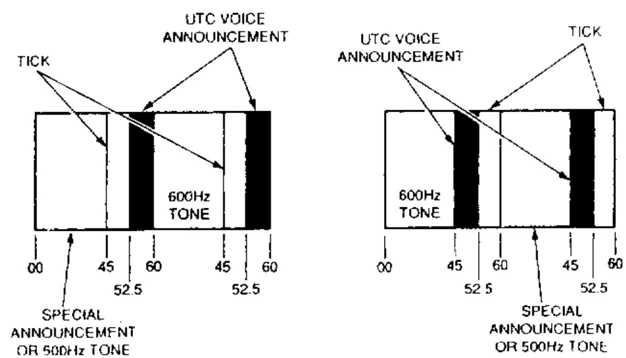
WILLIAM SHEETS *Fig. 76-27*

L1 and L2 are wound on 4" diameter 10" form and are 200 turns of #24 wire. PVC pipe can be used.

WWV RECEIVER



WWV and WWVH minute signals



ELECTRONICS NOW *Fig. 76-28*

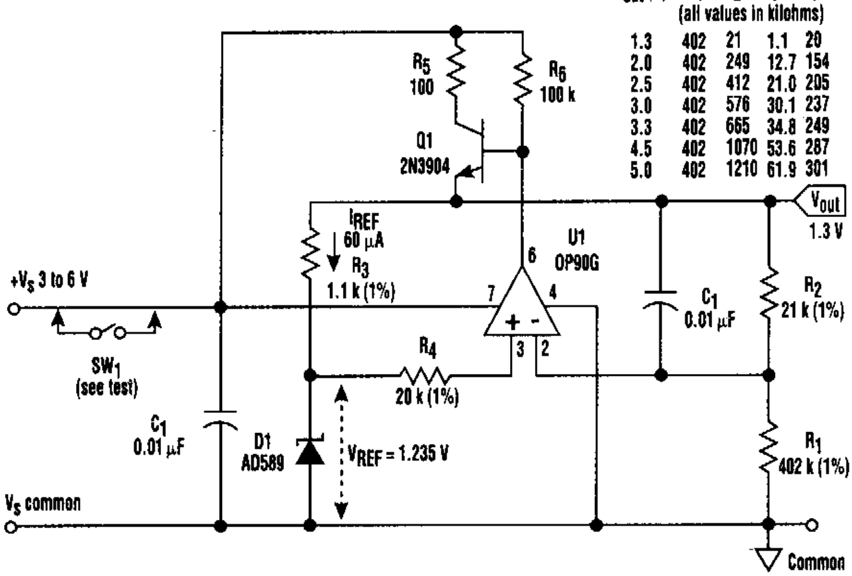
This receiver for 10-MHz WWV signals uses a 10.7-MHz FM receiver IF transformers as front-end components. It is a superhet with a 455-kHz IF frequency. By changing the front-end components 5- or 15-MHz reception could be obtained. A 3- to 6-foot antenna is usually adequate.

77

Reference Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOW-VOLTAGE REFERENCE



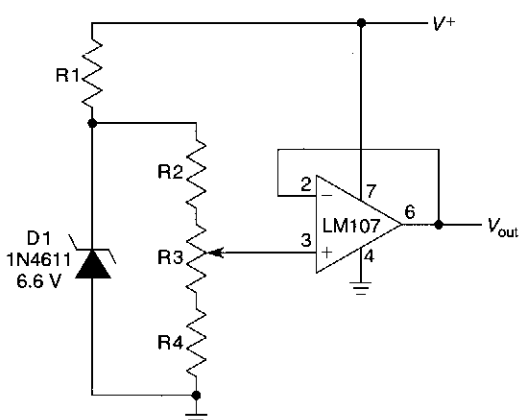
Output table

Vout (V)	R1	R2	R3	R4
	(all values in kilohms)			
1.3	402	21	1.1	20
2.0	402	249	12.7	154
2.5	402	412	21.0	205
3.0	402	576	30.1	237
3.3	402	665	34.8	249
4.5	402	1070	53.6	287
5.0	402	1210	61.9	301

ELECTRONIC DESIGN *Fig. 77-1*

This circuit illustrates a number of techniques that are useful for low-voltage, series-mode, power-efficient references. Intended for output currents of up to 10 mA, this design has an enabled standby current of about 100 μA; it can be easily programmed over a wide range of output voltages.

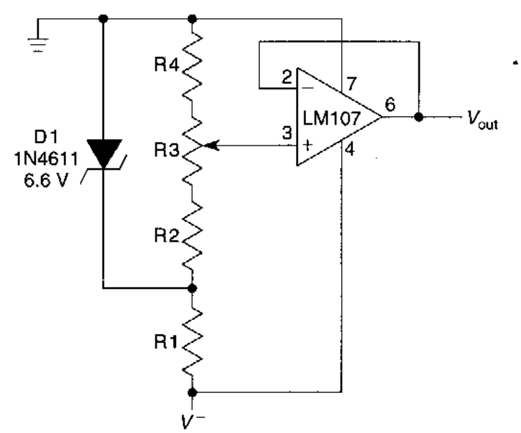
POSITIVE VOLTAGE REFERENCE



NATIONAL SEMICONDUCTOR *Fig. 77-2*

D1 is used as a reference. R2, R3, and R4 provide desired output voltage to the op-amp voltage follower.

NEGATIVE VOLTAGE REFERENCE



NATIONAL SEMICONDUCTOR *Fig. 77-3*

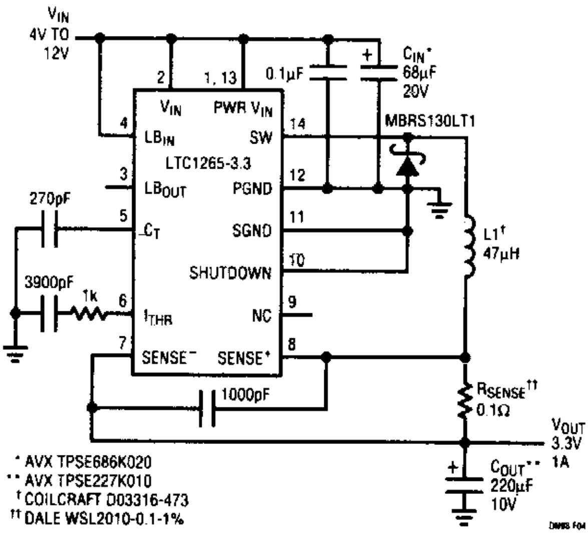
D1 is used as a reference. R2, R3, and R4 are voltage dividers to obtain desired output voltage to the op-amp voltage follower.

78

Regulator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

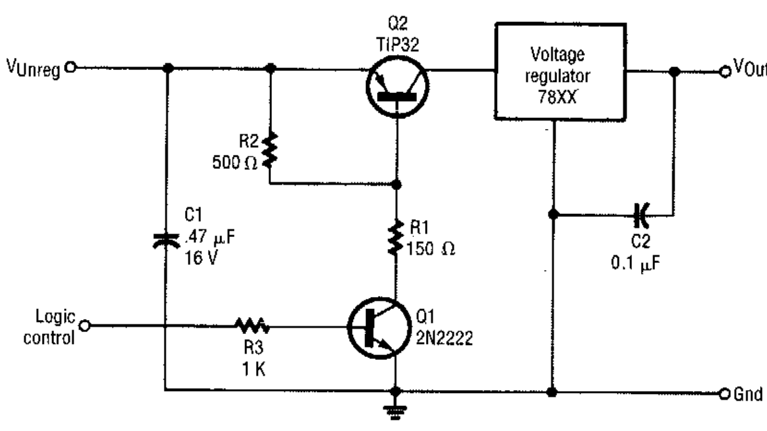
3.3-V 1-A SURFACE-MOUNT REGULATOR



LINEAR TECHNOLOGY ***Fig. 78-1***

This figure shows a typical LTC1265 surface-mount application. It provides 3.3 V at 1 A from an input voltage range of 4 V to 12 V. The peak efficiency approaches 93% at mid-current levels.

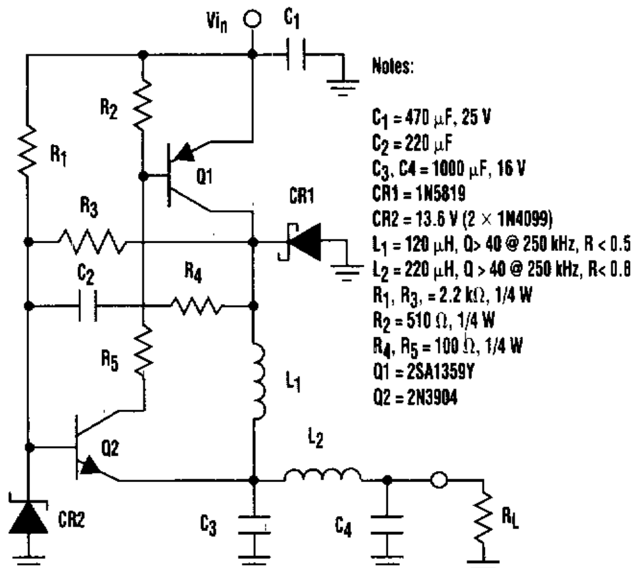
LOGIC CONTROL OF 78XX REGULATOR



ELECTRONICS NOW ***Fig. 78-2***

Transistors can be used to control any 78xx series regulator with logic signals. Both transistors are controlled by the logic level present at the base of Q1.

LOW-COST STEP-DOWN REGULATOR



Input = 18.0 V, Output taken at C3

Output (V)	Load (Ohm)	Ripple (mVpp)	Ripple (kHz)	Eff. (%)
12.54	1k	50	4	67
12.52	90.9	25	112	86.5
12.49	47.6	40	58	88.9
12.45	24.4	70	31	91.8

Input = 18.0 V, Output taken at C4

Output (V)	Load (Ohm)	Ripple (mVpp)	Ripple (kHz)	Eff. (%)
12.53	1k	58	.08	67
12.46	90.9	1.5	—	86.5
12.37	47.6	1.5	—	88.4
12.20	24.4	1.5	—	90.4

Output taken at C3, Load = 24.4 Ohms

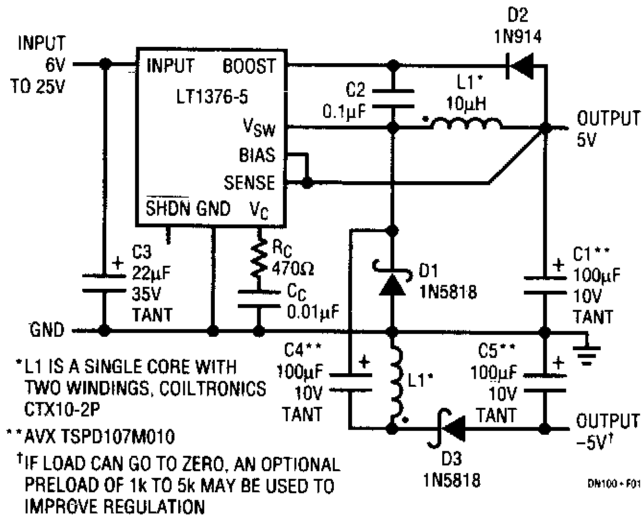
Input (V)	Output (V)	Ripple (mVPP)	Ripple (kHz)	Eff. (%)
15.0	12.35	73	17.8	93.4
18.0	12.45	70	31.0	91.8
21.0	12.53	75	43.3	90.8

ELECTRONIC DESIGN

Fig. 78-3

This inexpensive and efficient discrete step-down regulator is based on a complementary transistor arrangement that uses both positive and negative feedback and is referenced to a Zener diode. Inductor L1 is selected to maintain the switching frequency above the audible range for the intended operating load. The output filter L2 and C4 reduces ripple to less than 10 mV p-p over a large range of loads, with only a slight decrease in efficiency.

DUAL-OUTPUT REGULATOR

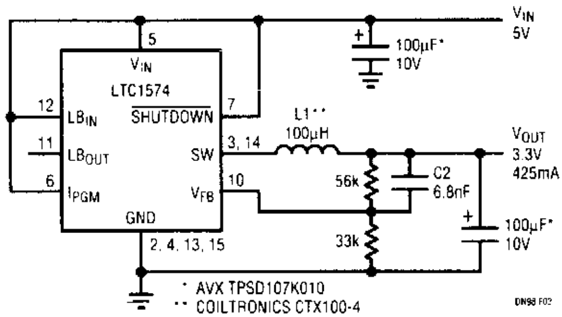


LINEAR TECHNOLOGY ***Fig. 78-4***

If load can go to zero, an optional preload of 1 to 5 kΩ can be used to improve regulation. Many modern circuit designs still need a dual polarity supply. Communication and data acquisition are typical areas where both 5 V and –5 V are needed for some of the IC chips.

The current mode architecture and saturating switch design allow the LT1376 to deliver up to 1.5-A load current from the 8-pin SO package. L1 is a 10-µH surface-mount inductor from Coiltronics. The second winding is used to create a negative-output SEPIC (Single-Ended Primary Inductance Converter) topology using D3, C4, C5, and the second half of F1. This converter takes advantage of the fact that the switching signal driving L1 as a positive buck converter is already the correct amplitude for driving a –5-V SEPIC converter. During switch-off time, the voltage across L is equal to the 5-V output plus the forward voltage of D1. An identical voltage is generated in the second winding, which is connected to generate –5 V using D3 and C5. Without C4, this would be a simple flyback winding connection with modest regulation. The addition of C4 creates the SEPIC topology. Note that the voltages swing at both ends of C4 is theoretically identical—even without the capacitor. The undotted end of both windings goes to a zero ac voltage node, so the equal windings will have equal voltages at the opposing ends. Unfortunately, coupling between windings is never perfect, and load regulation at the negative output suffers as a result. The addition of C4 forces the winding potentials to be equal and gives much better regulation.

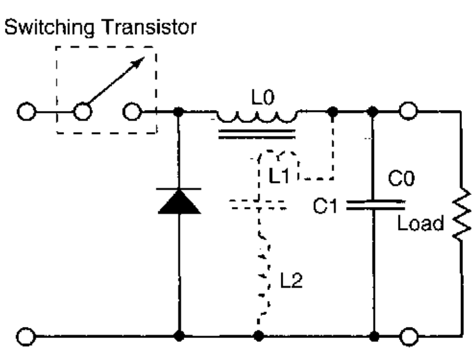
LOW-NOISE REGULATOR (5 TO 3.3 V)



LINEAR TECHNOLOGY *Fig. 78-5*

In some applications, it is important not to introduce any switching noise within the audio frequency range. To circumvent this problem, a feed-forward capacitor can be used to shift the noise spectrum up and out of the audio band with C2 being the feed-forward capacitor. The peak-to-peak output ripple is reduced to 30 mV over the entire load range. A toroidal surface mount inductor L1 is chosen for its excellent self-shielding properties.

REDUCING RIPPLE IN A SWITCHING VOLTAGE REGULATOR



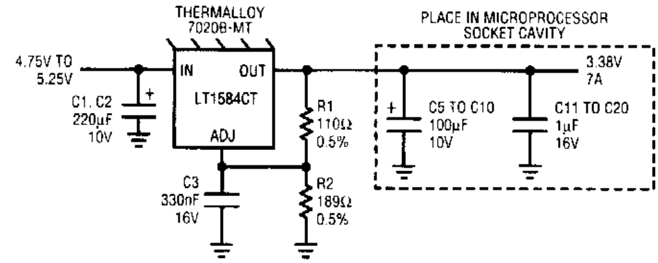
NASA TECH BRIEFS *Fig. 78-6*

Simple additional circuitry that consists of relatively small components can reduce the output ripple by a factor of about 10. The additional components are indicated by the dashed lines.

A current opposing the ripple is injected into the filter capacitor. The essence of the present technique is to inject, into this capacitor, a current opposite to that which already flows into this capacitor. A small additional winding, L1, in inductor L0 provides transformer coupling to generate the current that opposes the original ripple current. The circuit from L1 through C0 is completed by a small additional external inductor L2 and coupling capacitor C1.

LOW-DROPOUT THREE-TERMINAL REGULATORS FOR NEW MICROPROCESSOR APPLICATIONS

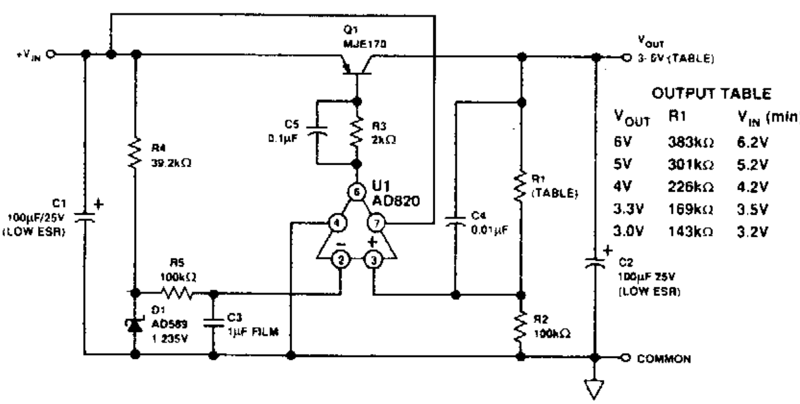
Recommended LT1584 Adjustable Circuit for the Intel P54CT Microprocessor



LINEAR TECHNOLOGY *Fig. 78-7*

The LT1584/LT1585/LTL1587 are high-performance, low-dropout regulators designed to meet the demands of the newest high speed, low voltage microprocessors. These devices are designed to regulate from 5-V supplies to output voltages between 1.25 V and 3.6 V. The LT1584 can provide up to 7 A of current, making it ideal for powerful Pentium processor or similar applications. The LT1585 can supply up to 4 A, while the LT1587 supplies up to 3 A. The excellent transient response capability allows them to maintain good regulation even with significant load steps. Fixed 3.3 V, 3.45 V, 3.6 V and adjustable output voltages are available.

LOW-DROPOUT REGULATOR



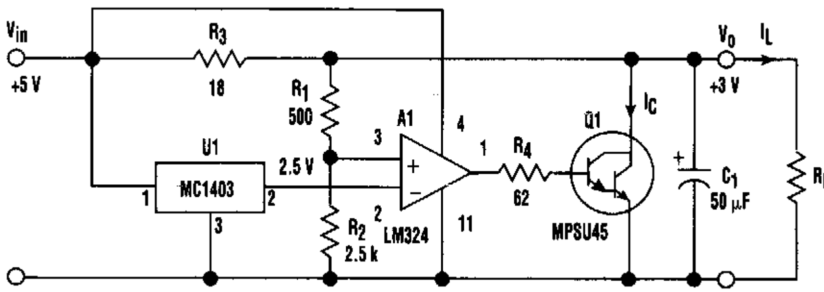
OUTPUT TABLE

V_{OUT}	R1	V_{IN} (min)
6V	383kΩ	6.2V
5V	301kΩ	5.2V
4V	226kΩ	4.2V
3.3V	169kΩ	3.5V
3.0V	143kΩ	3.2V

ELECTRONIC DESIGN *Fig. 78-8*

This low-dropout reference produces a 4.5-V output from a supply just a few hundred mV greater. With 1-mA dc loading, it maintains a stable 4.5-V output for inputs down to 4.7 V.

POSITIVE REGULATOR SINKS CURRENT



ELECTRONIC DESIGN *Fig. 78-9*

Generally speaking, conventional positive voltage regulators can only source current; they can't sink it. However, the positive regulator shown breaks that rule because it can perform both functions. The idea is to have the control transistor Q1 in shunt so that the regulator can either source or sink current.

The circuit provides +3-V output from a +5-V supply. U1 is a bandgap reference that supplies a stable +2.5-V reference to the error amplifier (A1). The output voltage (V_0) is sampled by the resistor network (R1 and R2). If V_0 were to increase, A1 will drive the base of Q1 harder, increasing the collector current (I_c).

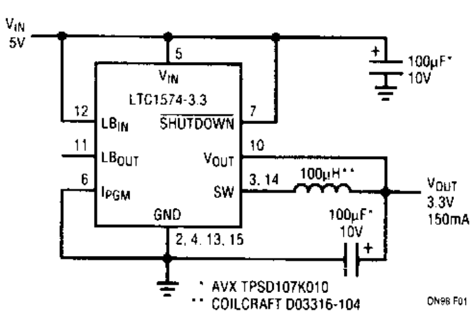
This increases the drop across R3 and V_0 decreases, thus regulating the output voltage. The output voltage is given by $V_0 = 2.5(1 + R_1/R_2)$.

Under no load conditions, Q1 draws 110 mA $[(V_{in} - V_0)/R_3]$. With a load connected, and as the regulator begins to source load current (I_L), I_c decreases to keep the drop across R3 constant.

At I_L = 100 mA, Q1 carries 10 mA. If R_L is connected to the positive supply higher than V_0, then the regulator must sink current, and I_L becomes negative. At I_L = –100 mA, Q1 carries 210 mA while maintaining the output voltage at +3 V. The output voltage will remain constant at +3 V—even if the load current changes sign.

With the proper heatsink on Q1, the regulator can sink more than 300 mA. If a "sink only" option is desired, the dissipation in Q1 can be reduced by using a 180-Ω resistor for R3. R4 limits the base current drive for Q1 and prevents the output of A1 from being clamped at 2 V_{BE}.

5- TO 3.3-V SURFACE-MOUNT SWITCHING REGULATOR



This converter provides 3.3 V at 150 mA from an input voltage of 5 V. Peak inductor current is limited to 340 mA by connecting pin 6 (I_{PGM}) to ground. For applications requiring higher output current, connect pin 6 to V_{in}. Under this condition, the maximum load current is increased to 425 mA.

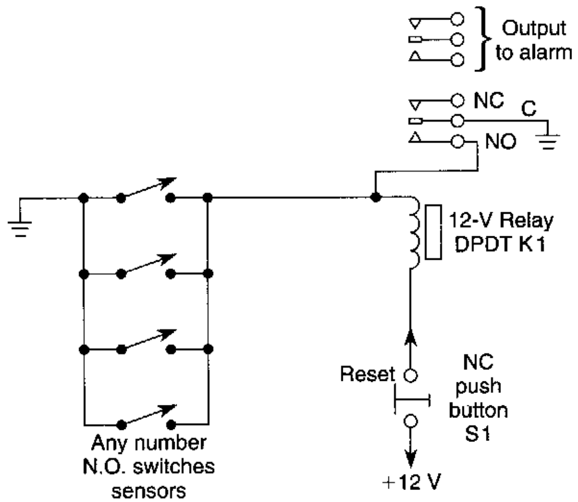
LINEAR TECHNOLOGY *Fig. 78-10*

79

Relay Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

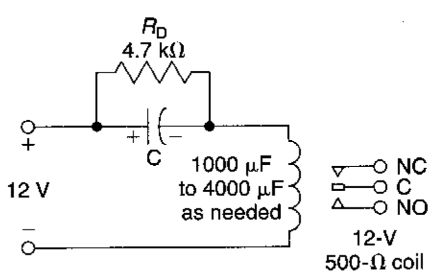
LATCHING RELAY ALARM CIRCUIT



WILLIAM SHEETS *Fig. 79-1*

Momentarily closing any sensors will cause K1 to latch. S1 must be depressed to reset circuit. If any sensor is still closed circuit will not reset.

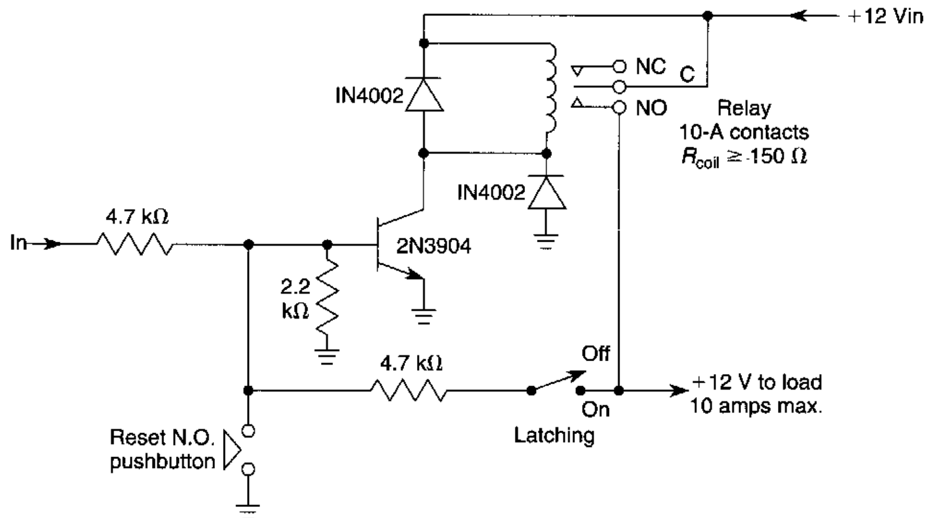
MOMENTARY RELAY CIRCUIT



WILLIAM SHEETS *Fig. 79-2*

The charging current of a capacitor can be used if a momentary relay-on circuit is needed. Depending on the relay characteristics, C will vary from 1000 to 4000 μF or so for a 1-s hold time if a 500-Ω relay is used. R_D discharges capacitor C to ready the circuit for the next operation. The value should be high enough so as not to maintain the relay closure at highest expected supply voltage.

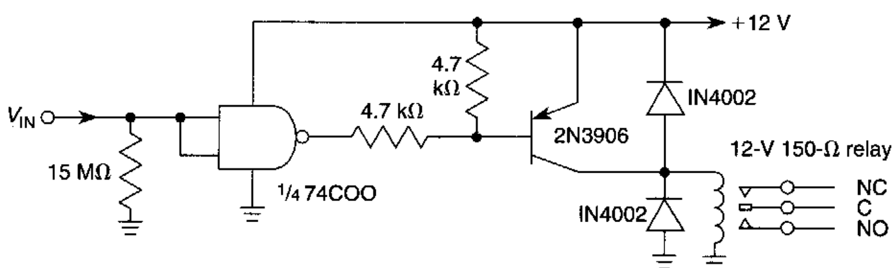
LATCHING RELAY DRIVER FOR +12-V LOADS



WILLIAM SHEETS **_Fig. 79-3_**

A 4-V signal will cause the relay to pull in when Q1 turns on. Latching is obtained by feedback through a 4.7-kΩ resistor. A switch is used to select latching or nonlatching operation. A NO pushbutton releases the circuit.

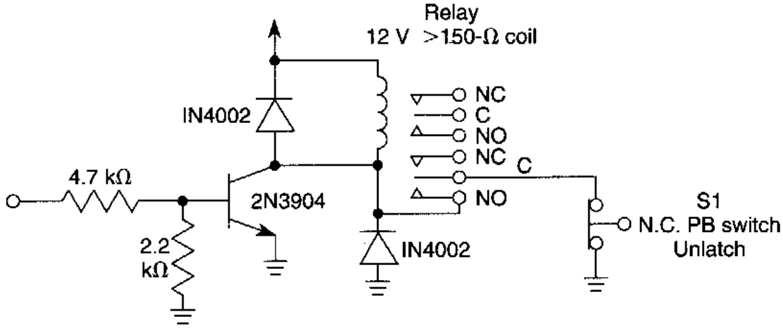
HIGH-IMPEDANCE RELAY DRIVER



WILLIAM SHEETS **_Fig. 79-4_**

A CMOS gate is used to drive a switching transistor and relay.

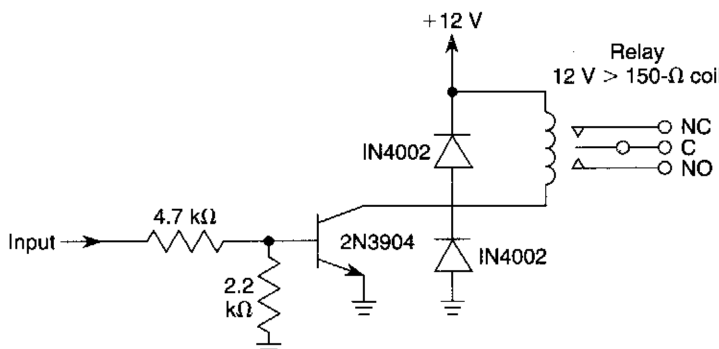
LATCHING RELAY DRIVER



WILLIAM SHEETS *Fig. 79-5*

An input of 4 V or greater will drive this circuit. When the relay pulls in, one pair of contacts is used to latch the relay closed. It will remain closed until S1 is pressed.

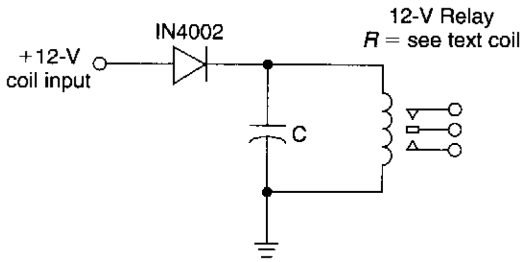
TRANSISTOR RELAY DRIVER



WILLIAM SHEETS *Fig. 79-6*

An input of 4 V or greater will drive this relay circuit.

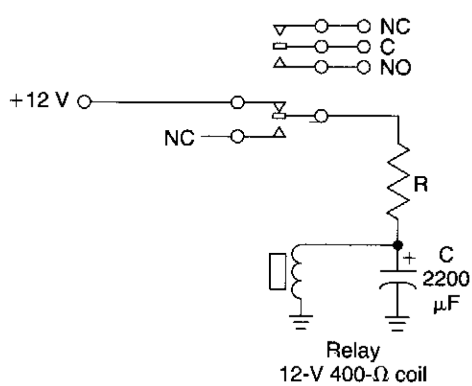
FAST TURN-ON/DELAYED-OFF RELAY CIRCUIT



WILLIAM SHEETS **Fig. 79-7**

C is a large capacitor that has a charge time of R_{supply} C, assuming $R_{supply} < R_{coil}$. The discharge time will be R_{coil} C neglecting relay coil inductance. With C = 10,000 μF and R_{coil} = 500 Ω, a release time constant of 5 seconds might be obtained. Many relays will hold in until the coil current decays to 25% of the pull-in current so that the actual time constant depends on the relay holding current.

LOW-FREQUENCY RELAY OSCILLATOR



WILLIAM SHEETS **Fig. 79-8**

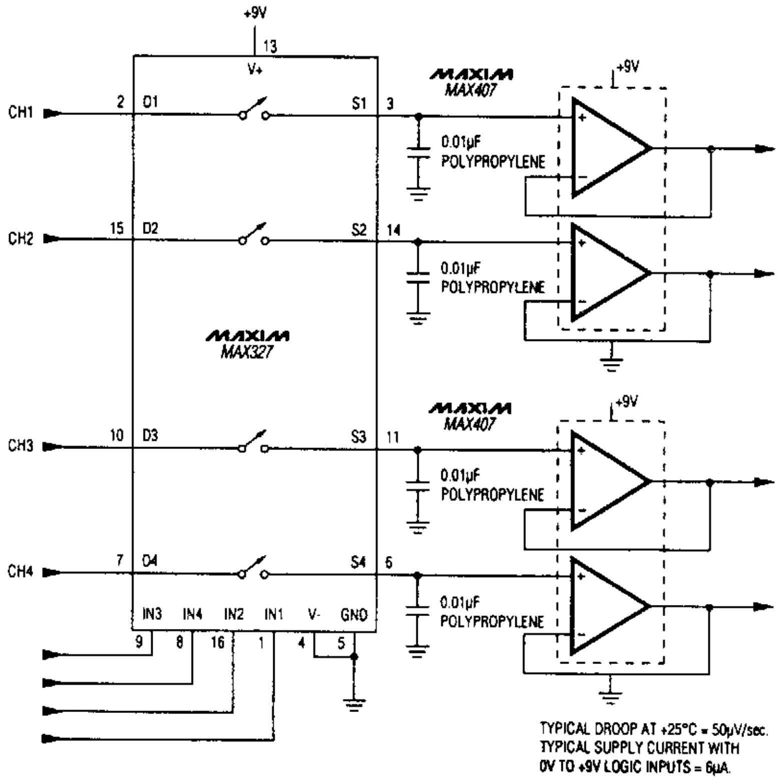
Depending on the value of C and the resistance of the relay coil, and the difference in pull-in and drop-out voltage, this circuit will oscillate at a low frequency. R limits in rush current to capacitor C to a level that the relay contacts can handle. Typically, for a 400-Ω relay, R can be 20 to 440 ohms. Flash rate is approximately 1 cycle/second, depending on the relay.

80

Sample-and-Hold Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

MICROPOWER 4-CHANNEL SAMPLE-AND-HOLD CIRCUIT

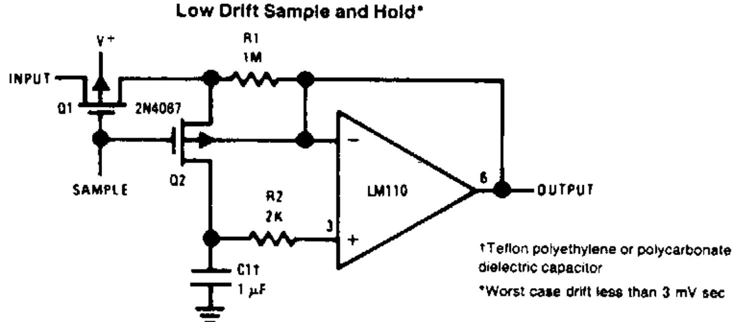


MAXIM *Fig. 80-1*

Three Maxim ICs make up this sample-and-hold circuit. The supply current is only 6 µA.

LOW-DRIFT SAMPLE AND HOLD

Low Drift Sample and Hold*



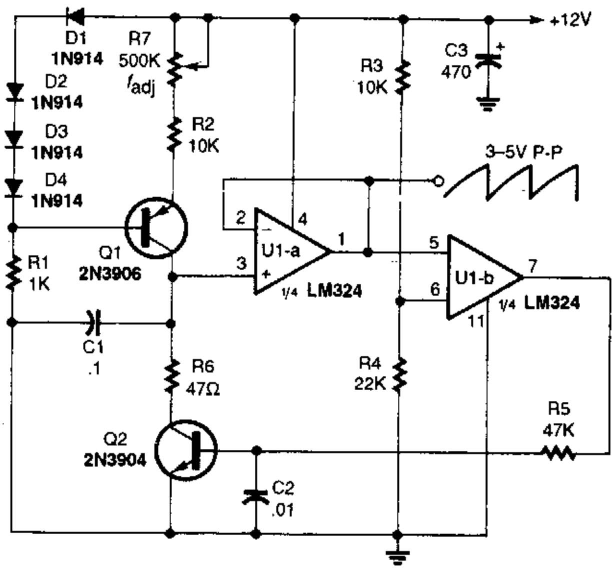
NATIONAL SEMICONDUCTOR *Fig. 80-2*

81

Sawtooth Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

OP-AMP LINEAR SAWTOOTH GENERATOR



POPULAR ELECTRONICS *Fig. 81-1*

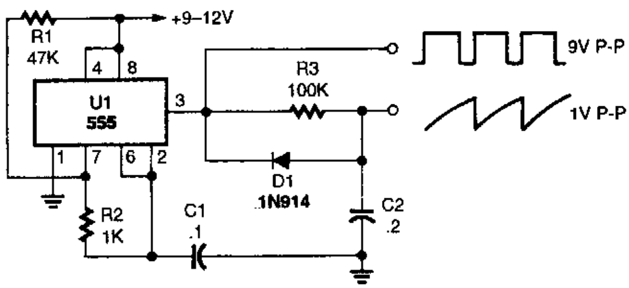
Q1 is connected in a simple constant-current generator circuit. The value of Q1's emitter resistor sets the constant-current level flowing from the transistor's collector to the charging capacitor, C1.

One op amp of an LM324 quad op-amp IC, U1-a, is connected in a voltage-follower circuit. The input impedance on the voltage follower is very high and offers little or no load on the charging circuit. The follower's output is connected to the input of U1-b, which is configured as a voltage comparator. The comparator's other input is tied to a voltage-divider setting the input level to about 8 V.

The output of U1-b at pin 7 switches high when the voltage at its positive input, pin 5, goes above 8 V. That turns on Q2, discharging C1. The sawtooth cycle is repeated over and over as long as power is applied to the circuit.

The sawtooth's frequency is determined by the value of C_1 and the charging current supplied to that capacitor. As the charging current increases, the frequency also increases, and vice versa. To increase the generator's frequency range, decrease the value of C_1, and to lower the frequency, increase the value of C_1. The output is about 3 to 5 V.

SAWTOOTH GENERATOR



POPULAR ELECTRONICS *Fig. 81-2*

A sawtooth waveform generator circuit using a 555 IC is shown. The IC is connected in an astable oscillator circuit with the majority of the output contained in the positive portion of the cycle. The negative output is a very brief pulse.

Capacitor C2 charges through R3 in a positive direction during the time that the IC's output (at pin 3) is high. When the output goes negative, C2 is rapidly discharged through D1 and the IC's output.

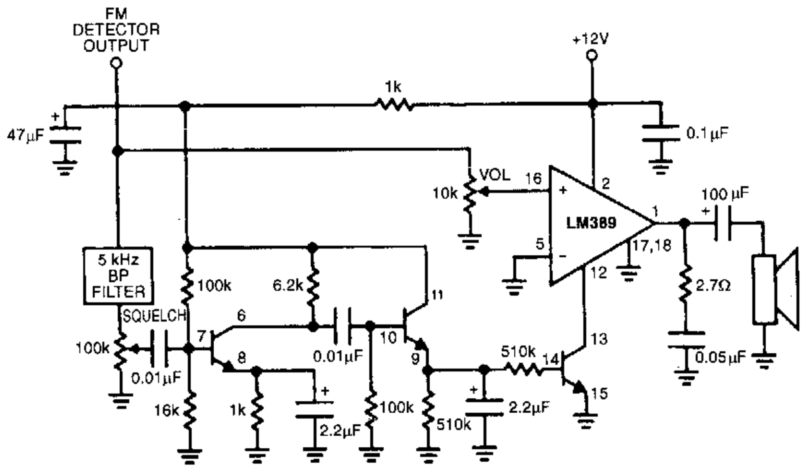
Peak-to-peak sawtooth output is about 1 V. The linearity of this circuit is best when R3 is as large as possible. The oscillator's frequency is about 200 Hz and can be increased by lowering either the value of R_1 or C_1; to decrease the frequency, increase the values of those components.

82

Scanner Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

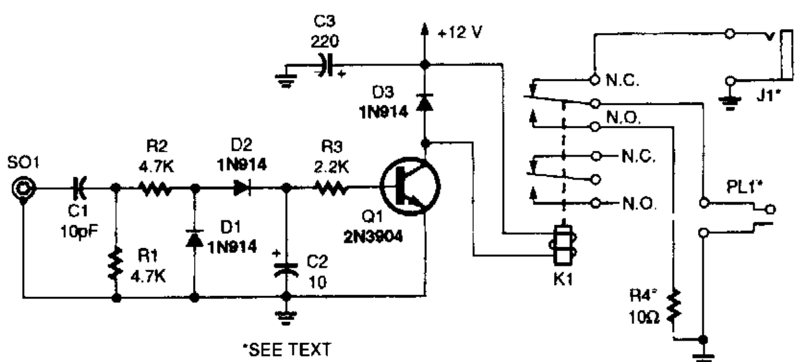
FM SCANNER NOISE SQUELCH



NATIONAL SEMICONDUCTOR *Fig. 82-1*

The LM389 is operated in the cut-off mode with pin 12 grounded via one of the internal transistors. A sample of detected noise is taken through a 5-kHz filter. Upon reception of signal, the detector output quiets, and noise level drops. This increases impedance at pin 12 of the LM389, causing audio to be passed. The three transistors are part of the LM389.

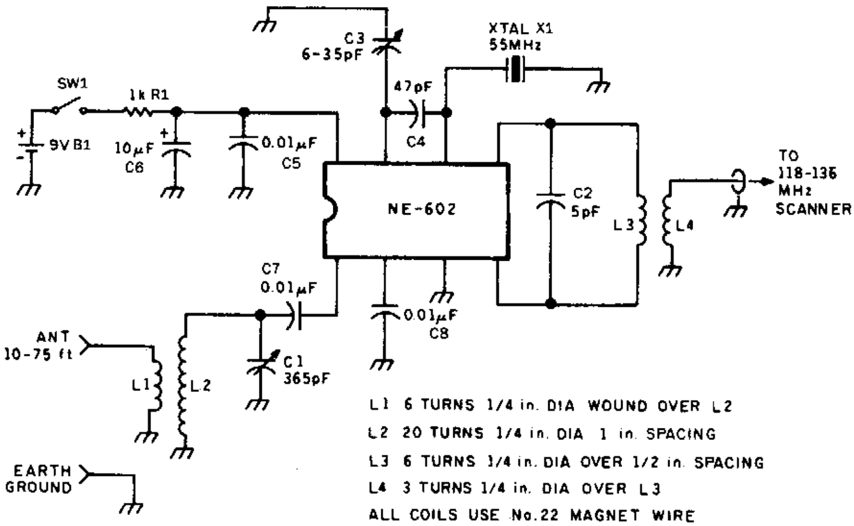
SCANNER SILENCER



POPULAR ELECTRONICS *Fig. 82-2*

When a scanner is used at amateur radio or CB stations, the scanner sometimes picks up transmitted signal and howls or squeals. When RF is detected, Q1 turns on, energizing K1 and disconnecting the scanner speakers. SO1 is connected to the transmit antenna lead via a tee fitting. C1 is optimum for 5- to 10-W 30-MHz use. For higher power or higher frequencies, reduce C1 to as low as needed. If C1 is so small as to be impractical, R1 can be shunted with a 10- or 22-pF capacitor, as needed.

SHORTWAVE CONVERTERS FOR SCANNERS



73 AMATEUR RADIO TODAY ***Fig. 82-3***

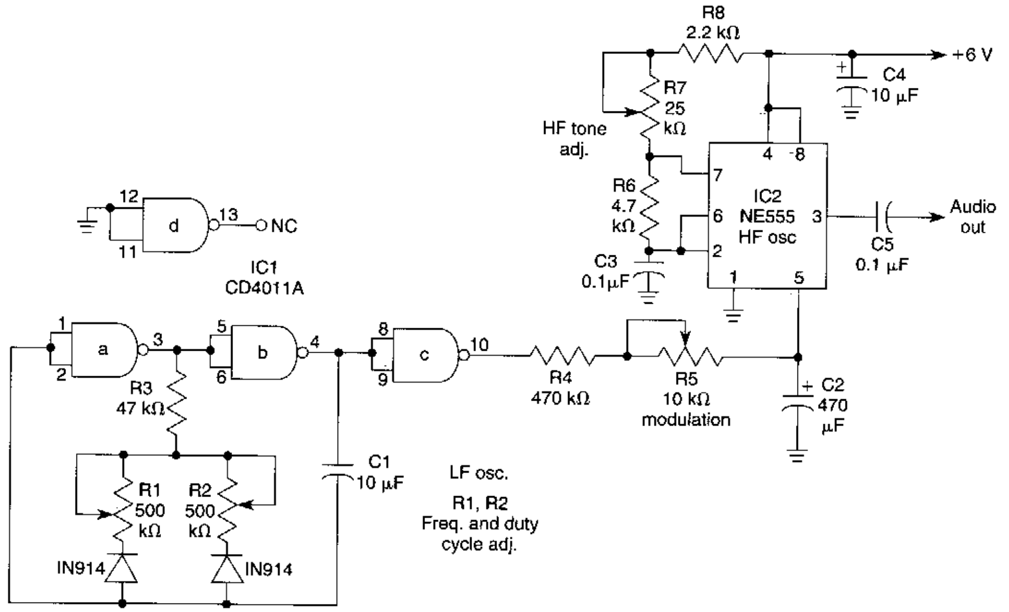
The AM aircraft band at 118 to 136 MHz is used in this converter design as an IF output. The second harmonic of the 55-MHz crystal (110 MHz) mixes with the shortwave input of 8 to 36 MHz. An NE602 IC is used for the mixer. Sensitivity is about 3 μV. If desired, a crystal tuning circuit for fine tuning can be obtained using a varactor.

83

Siren, Warbler, and Wailer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

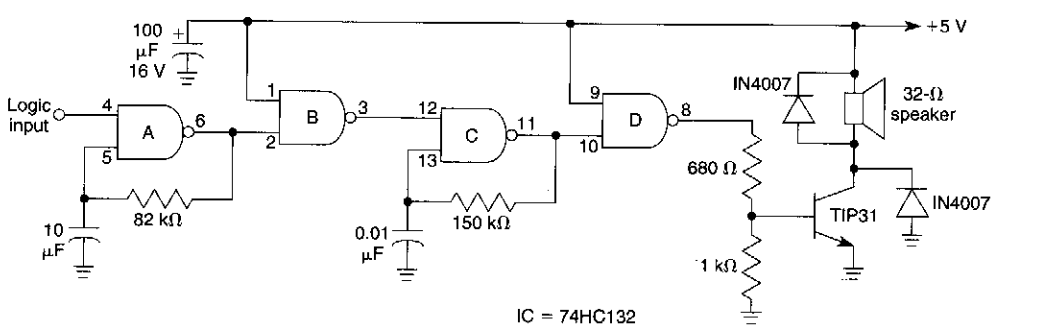
FIRE SIREN



WILLIAM SHEETS *Fig. 83-1*

IC1 is an LF oscillator that is variable in attack and decay time with R1 and R2. The LF output modulates HF oscillator IC2. R5 varies the modulation depth. By proper control adjustment, sirens of various types can be simulated.

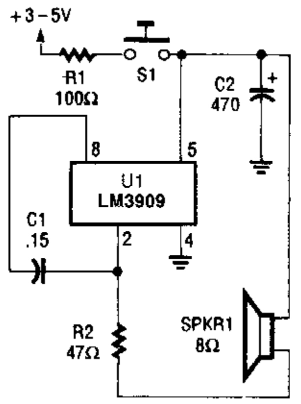
WARBLE OSCILLATOR



WILLIAM SHEETS *Fig. 83-2*

Sections A & B form an oscillator running at 2 Hz, which gates sections C and D, a 1-kHz oscillator. This drives the TIP31 speaker driver.

ELECTRONIC SIREN

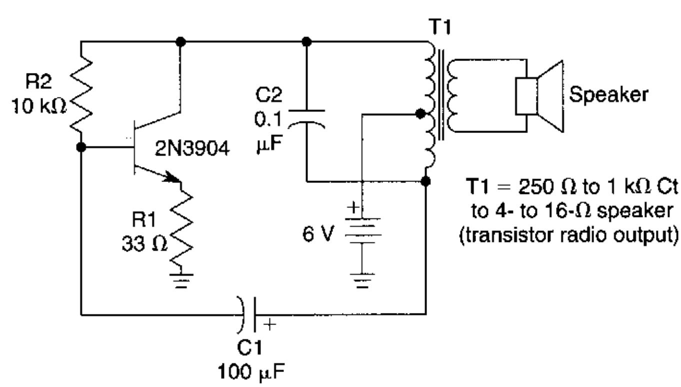


POPULAR ELECTRONICS *Fig. 83-3*

In this circuit, the LM3909 is used in a simple electronic siren. When S1 is closed, C2 begins to charge rapidly through R1. When the charge on C2 reaches about 1 V, the oscillator starts. As the voltage across C1 increases toward $+V$, the oscillator's output frequency also increases. Releasing (opening) S1 removes power from the circuit. The oscillator continues to operate, with a decline in output volume and frequency until C1 discharges to about the 1-V level.

Experiment with the siren circuit by selecting different R_1/C_2 combinations to obtain a desired rise and fall output. Change the value of C_1 to vary the oscillator's frequency. Keep the value of R_2 at or above 47 Ω to protect the IC from drawing too much current.

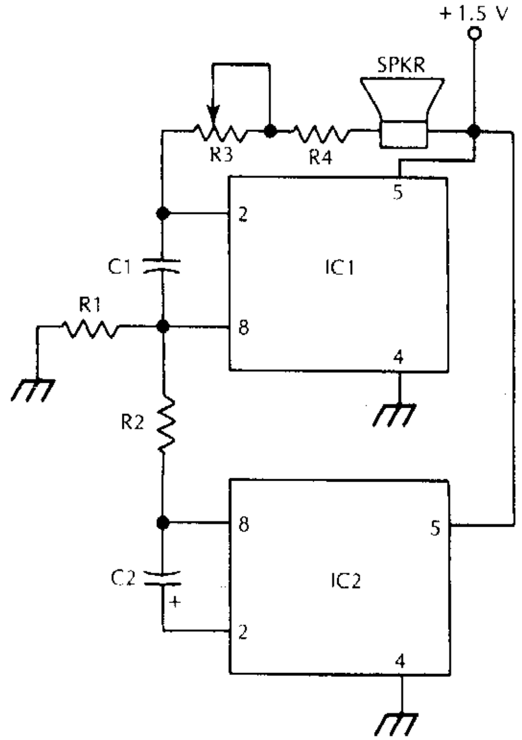
WAILING SOUND GENERATOR



WILLIAM SHEETS *Fig. 83-4*

In this circuit, C2 and T1 determine the tone generated and C_1/R_2 control the blocking rate. The signal produced is an interrupted tone, like a police whistle or toy ray gun, depending on C1 and C2.

TWO-TONE SIREN



IC1, IC2	LM3909 LED flasher/oscillator
SPKR	small 8 Ω speaker
C1	0.1 μF capacitor
C2	470 μF 6 V electrolytic capacitor
R1	10 kΩ ¼ W 5% resistor
R2	4.7 kΩ ¼ W 5% resistor
R3	100 Ω potentiometer
R4	33 Ω ¼ W 5% resistor

McGRAW-HILL ***Fig. 83-5***

IC1 generates the main siren tone while IC2 generates a low-frequency square wave, switching IC1 between two different tones.

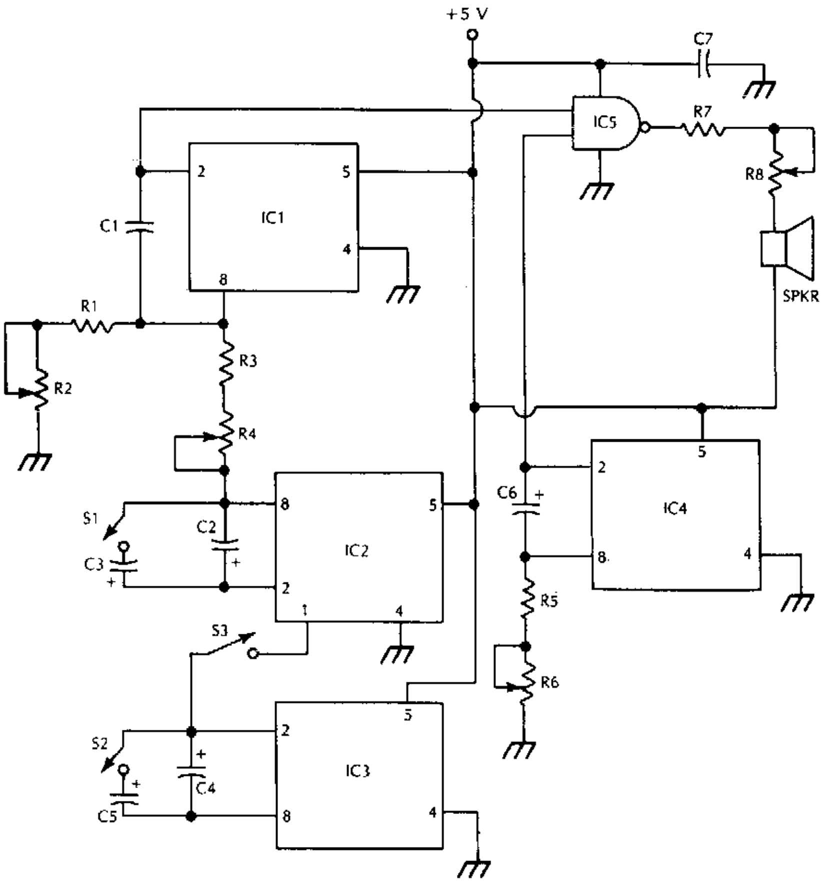
84

Sound-Effects Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

Complex Sound-Effect Generator
Dual-Tone Generator
Surf Man Sound Generator
Electronic Whistle
Bird-Chirp Sound-Effect Generator
Robotic Chatter Sound Generator
Electronic Wind Chime
Gunshot Sound-Effects Generator

COMPLEX SOUND-EFFECT GENERATOR

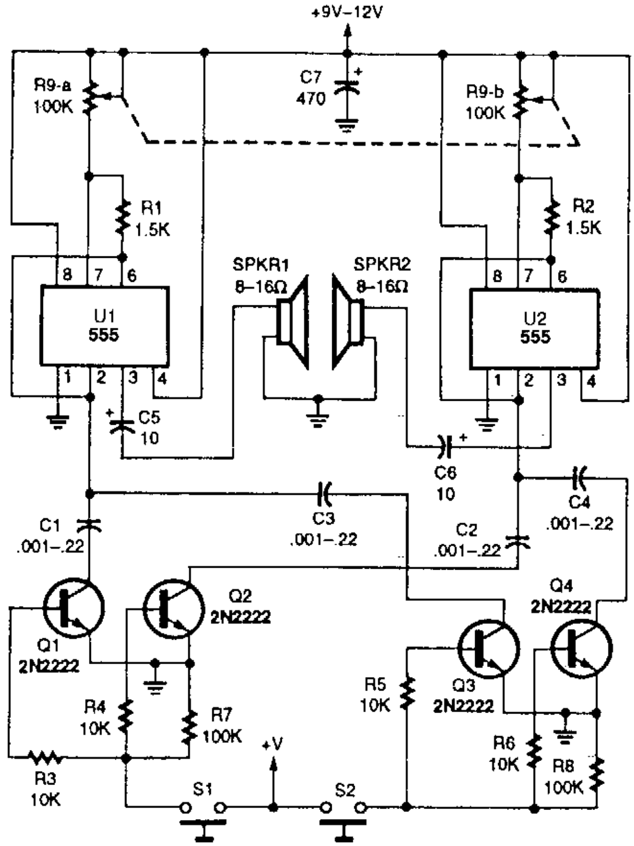


IC1–IC4	LM3909 LED flasher/oscillator	C4	47 μF 6 V electrolytic capacitor
IC5	7400 quad NAND gate	C5	33 μF 6 V electrolytic capacitor
SPKR	small loudspeaker	C6	4.7 μF 6 V electrolytic capacitor
S1, S2, S3	SPST switch	R1, R3, R5	2.2 kΩ ¼ W 5% resistor
C1	0.1 μF capacitor	R2, R4, R6	10 kΩ potentiometer
C2	22 μF 6 V electrolytic capacitor	R7	33 Ω ¼ W 5% resistor
C3	10 μF 6 V electrolytic capacitor	R8	100 Ω potentiometer

McGRAW-HILL ***Fig. 84-1***

This system uses four free running oscillators to produce a wide variety of complex sounds. LF oscillator IC3 modulates IC2, which modulates IC1. The audio from IC1 is combined with a variable frequency from IC4. Switches at various points allow oscillators IC3 to be switched in or out, IC1 and IC2 to be varied in frequency, and IC4 also can be varied in frequency. The circuit is not critical and different arrangements can be tried to produce various sound effects.

DUAL-TONE GENERATOR

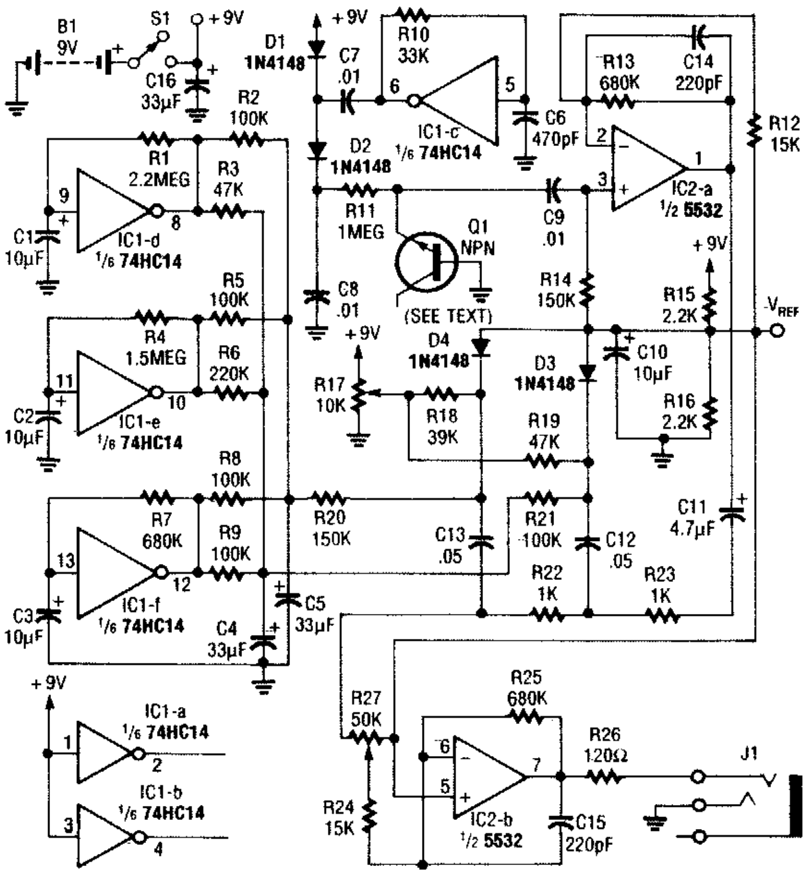


POPULAR ELECTRONICS *Fig. 84-2*

Two 555 oscillator/timers are configured similarly as audio oscillators, with each oscillator feeding a separate speaker.

A dual 100-kΩ potentiometer is use to tune the two oscillators simultaneously. The oscillators' frequency range is controlled by a dual-transistor switch, which selects the timing capacitor for both oscillators. Although the circuit only shows two range-switching circuits, any number can be added by simply duplicating the two-transistor switching circuit.

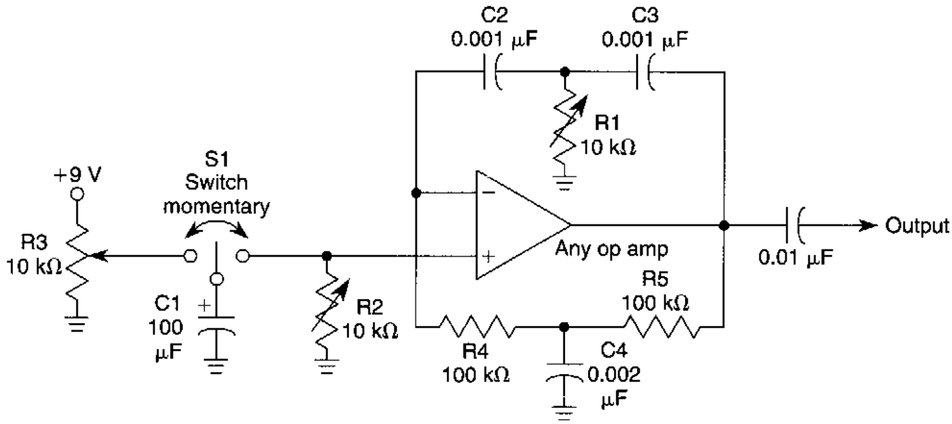
SURF MAN SOUND GENERATOR



ELECTRONICS NOW *Fig. 84-3*

Three low-frequency oscillators (IC1d, e, f) are used to simulate "wave action" of the surf. Q1 is an emitter-base junction used as a diode noise generator, biased by dc derived from oscillator IC1-c. The noise is fed into two voltage-controlled filters R22, R23, C12, C13, with D3 and D4 as "tuning" elements. The low-frequency oscillator signals randomly vary the filters, therefore, the spectrum of the noise signal fed through them. This simulates the sound of a surf.

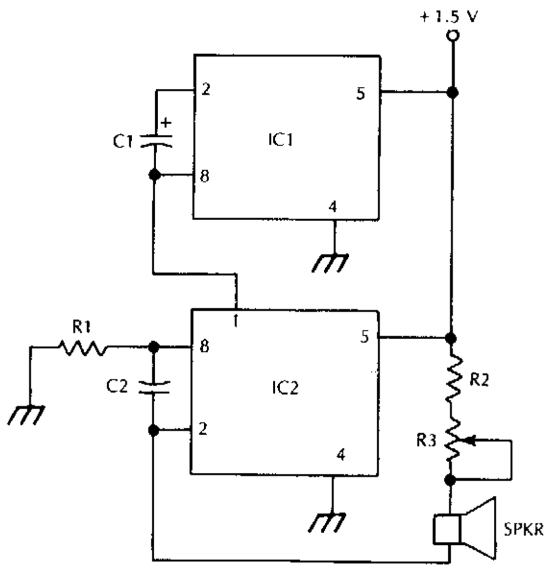
ELECTRONIC WHISTLE



WILLIAM SHEETS *Fig. 84-4*

The circuit shown is a twin-tee oscillator. R1 varies the pitch, R2 the duration, and R3 the format (bell, rise & fall time, etc.). Vary R4, R5, C4 and C2, C3 for large shifts in frequency.

BIRD-CHIRP SOUND-EFFECT GENERATOR

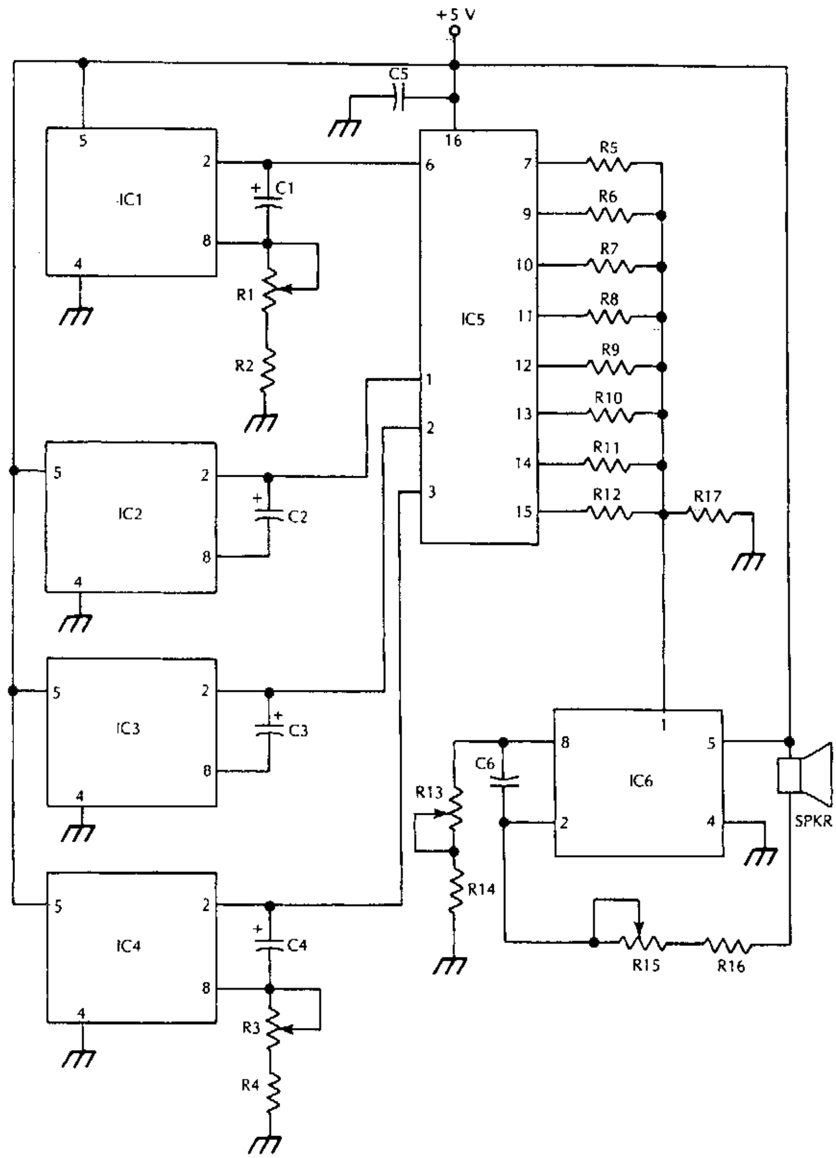


IC1, IC2	LM3909 LED flasher/oscillator IC
C1	33 μF 5 V electrolytic capacitor
C2	0.1 μF capacitor
R1	15 kΩ ¼ W 5% resistor
R2	33 Ω ¼ W 5% resistor
R3	250 Ω potentiometer
SPKR	small 8 Ω speaker

McGRAW-HILL *Fig. 84-5*

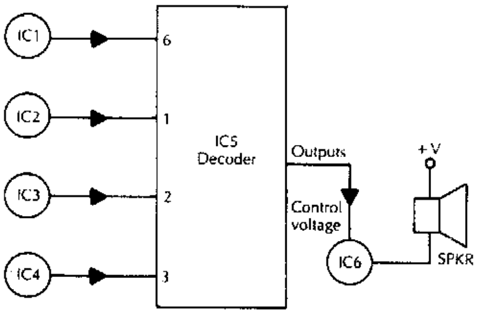
A low-frequency oscillator modulates a higher frequency oscillator, which drives the speaker.

ROBOTIC CHATTER SOUND GENERATOR



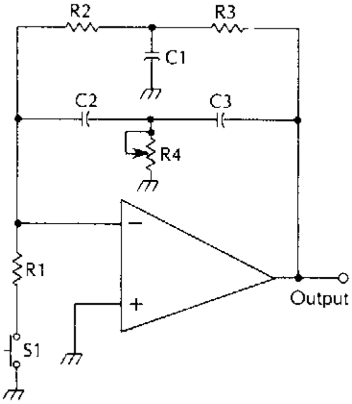
McGRAW-HILL *Fig. 84-6*

This circuit simulates sound effects of a robot, for toy or novelty applications.



IC1–IC4, IC6	LM3909 LED flasher/oscillator	R5	6.8 kΩ ¼ W 5% resistor
IC5	74LS138 three-line to eight-line decoder	R6	10 kΩ potentiometer
SPKR	small loudspeaker	R7	2.2 kΩ ¼ W 5% resistor
C1	10 µF 10 V electrolytic capacitor	R8	33 kΩ ¼ W 5% resistor
C2	22 µF 10 V electrolytic capacitor	R9	3.9 kΩ ¼ W 5% resistor
C3	33 µF 10 V electrolytic capacitor	R10	4.7 kΩ ¼ W 5% resistor
C4	100 µF 10 V electrolytic capacitor	R11	100 kΩ ¼ W 5% resistor
C5	0.01 µF capacitor	R12	470 kΩ ¼ W 5% resistor
C6	0.1 µF capacitor	R15	100 Ω potentiometer
R1, R3, R14	10 kΩ potentiometer	R16	33 Ω ¼ W 5% resistor
R2, R4, R13	3.3 kΩ ¼ W 5% resistor	R17	1 MΩ ¼ W fixed resistor

ELECTRONIC WIND CHIME

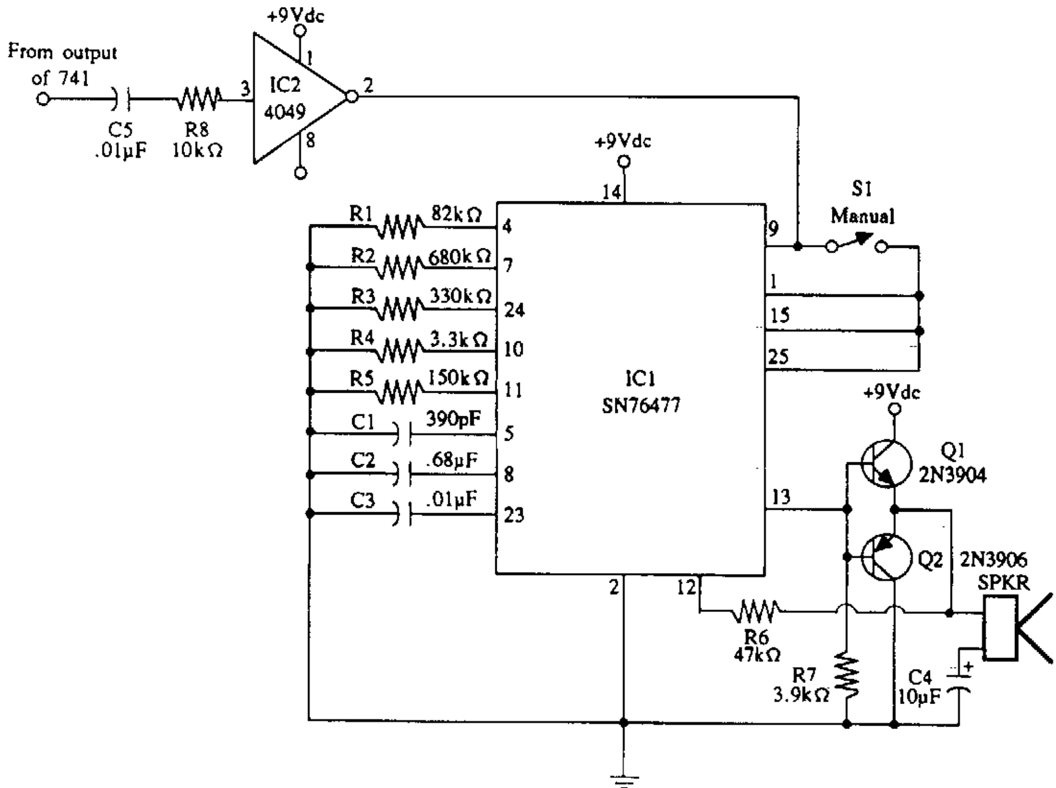


The value of R_4 controls the damping or decay time of the feedback circuit (a twin Tee oscillator). When S1 is closed, the circuit breaks into oscillation. When S1 is opened, the circuit stops oscillating generating a decaying tone like a bell. The frequency is approximately ½*RC*. C1, C2, and C3 are typically in the 0.01-µF range.

McGRAW-HILL

Fig. 84-7

GUNSHOT SOUND-EFFECTS GENERATOR



McGRAW-HILL *Fig. 84-8*

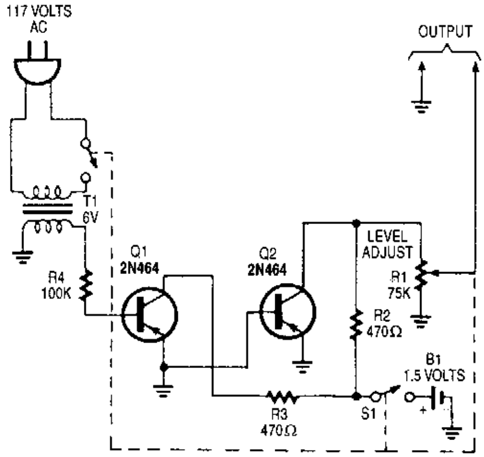
Gunshot sound-effects generator built around a Texas Instruments SN76477 sound chip. An input pulse causes IC1 to generate a gunshot sound.

85

Square-Wave Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

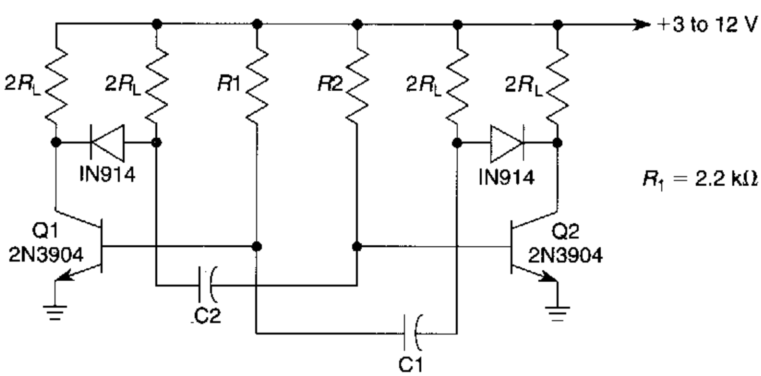
SQUARE-WAVE GENERATOR



POPULAR ELECTRONICS *Fig. 85-1*

A 60-Hz waveform from T1 drives an audio amplifier to clipping. Output is 60 Hz with about 0- to 1.4-V p-p amplitude.

SHARP SQUARE WAVEFORMS FROM MULTIVIBRATOR



$R_1 = 2.2\ \text{k}\Omega \qquad R_{L1} \geq \frac{R_1}{H_{FE}\,Q_1}$

WILLIAM SHEETS *Fig. 85-2*

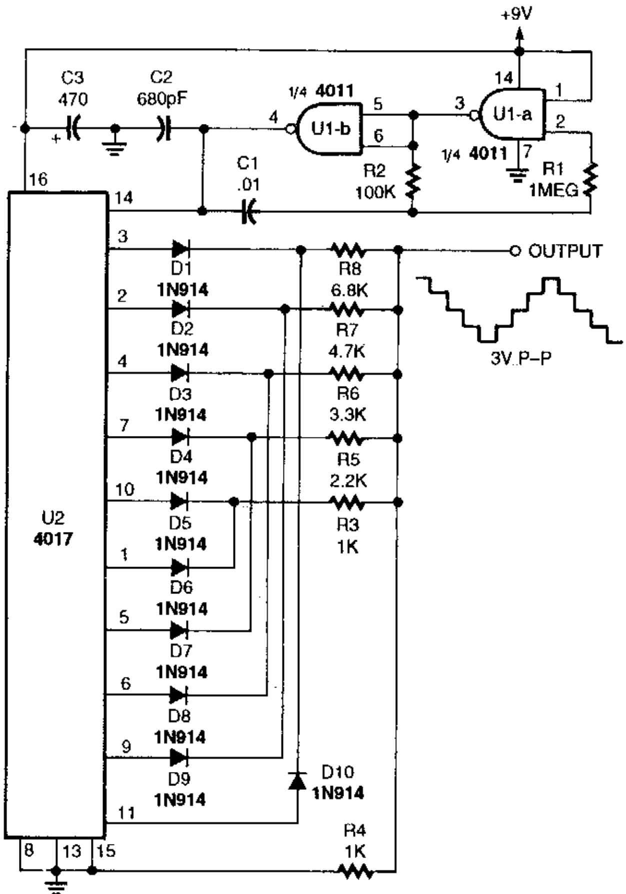
By using diodes as shown, the loading effect on the collector of the transistors caused by the timing capacitors can be avoided. As the collector of the transistors rises toward V_{CC}, the diode disconnects the timing capacitors.

86

Staircase Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

STEPPED TRIANGLE WAVEFORM GENERATOR

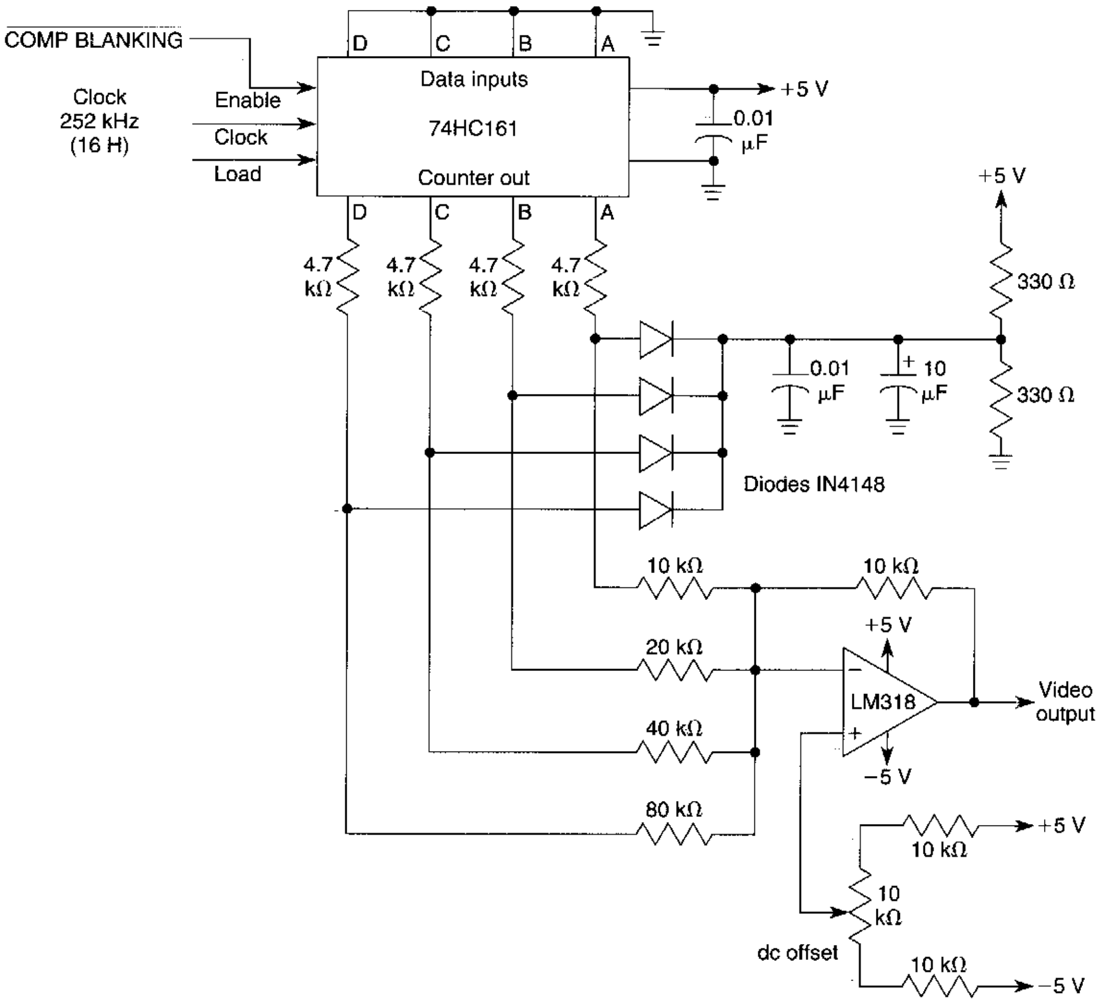


POPULAR ELECTRONICS *Fig. 86-1*

Two gates of a 4011 quad two-input NAND gate (U1) are connected in a pulse generator circuit. The square output at pin 4 of U1-b, connects to the clock input, pin 14, of a 4017 decade counter IC (U2). For each-input clock pulse, the 4017 takes a single step. Because the 4017 counter is set up to count ten and then repeat the count, the stepped output frequency will only be 1/10 of the clock frequency. For a 100-Hz output, the clock generator must operate at 1 kHz.

The 4017's positive output pulses begin at pin 3 and progress to pin 11 in a serial manner. The first output pulse, at pin 3, passes through D1 and R8 and appears across R4 to produce the first step up the triangle. The second pulse is routed through D2 and R7 to produce the second step. The outputs at pins 10 and 1 form the top of the waveform and outputs at pins 5, 6, 9, and 11 produce the down steps.

VIDEO STAIRCASE GENERATOR

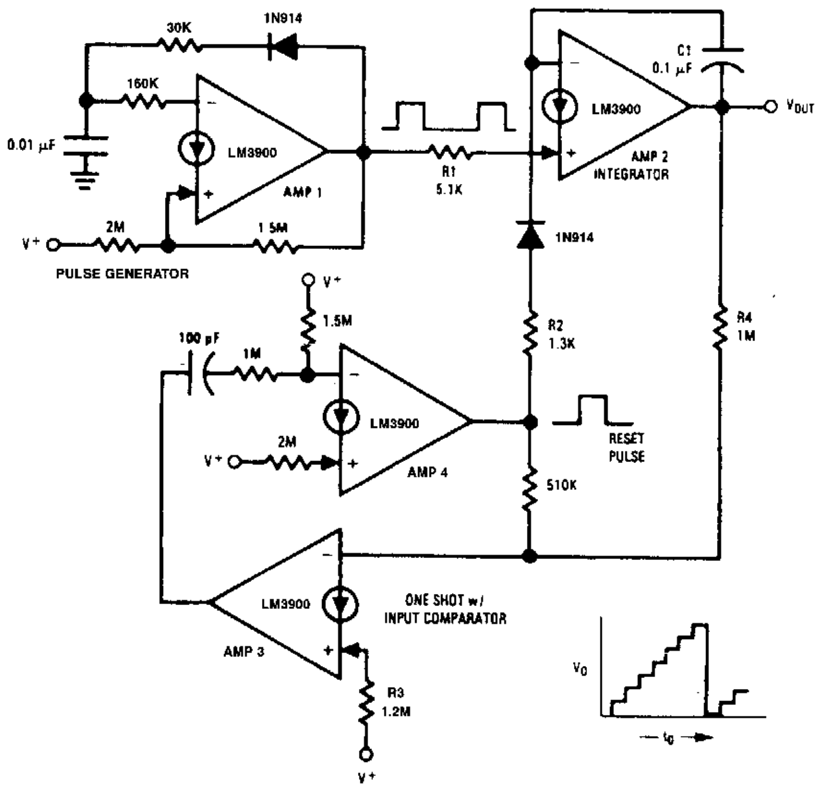


WILLIAM SHEETS

Fig. 86-2

Using a 74HC161 counter and a simple D-A converter using an op amp and resistor network, a very simple video staircase generator for gray-scale generation (12 bars) can be obtained. The output is clean and mostly free of "glitches."

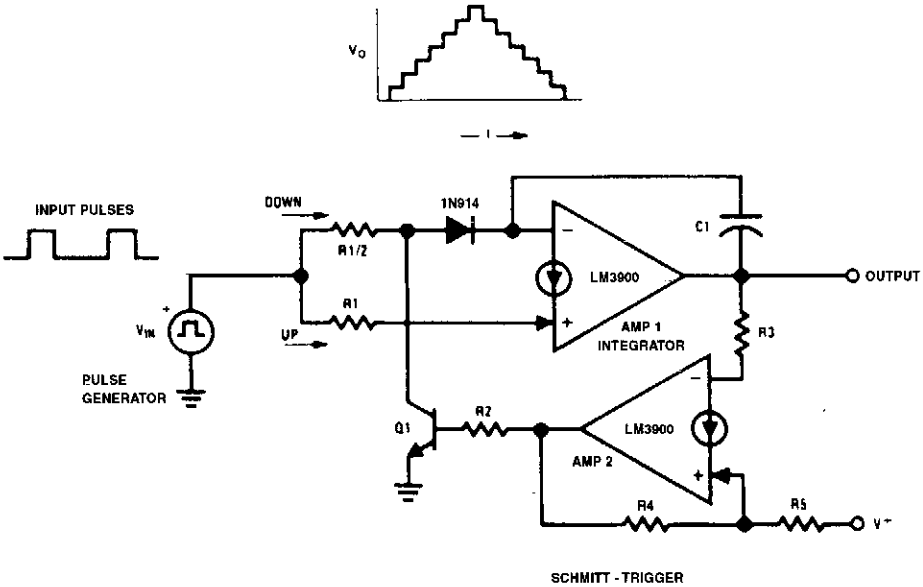
FREE-RUNNING STAIRCASE WAVE GENERATOR



NATIONAL SEMICONDUCTOR *Fig. 86-3*

This free-running staircase generator uses all four of the amplifiers, which are available in one LM3900 package. Amp 1 provides the input pulses that "pump up" the staircase via resistor R1. Amp 2 does the integrate and hold function and also supplies the output staircase waveform. Amps 3 and 4 provide both a compare and a one-shot multivibrator function. Resistor R4 is used to sample the staircase output voltage and to compare it with the power supply voltage (*V*+) via R3. When the output exceeds approximately 80% of *V*+ the connection of Amps 3 and 4 causes a 100-µs reset pulse to be generated. This is coupled to the integrator (AMP2) via R2 and causes the staircase output voltage to fall to approximately 0 V. The next pulse out of Amp 1 then starts a new stepping cycle.

UP/DOWN STAIRCASE WAVE GENERATOR



NATIONAL SEMICONDUCTOR ***Fig. 86-4***

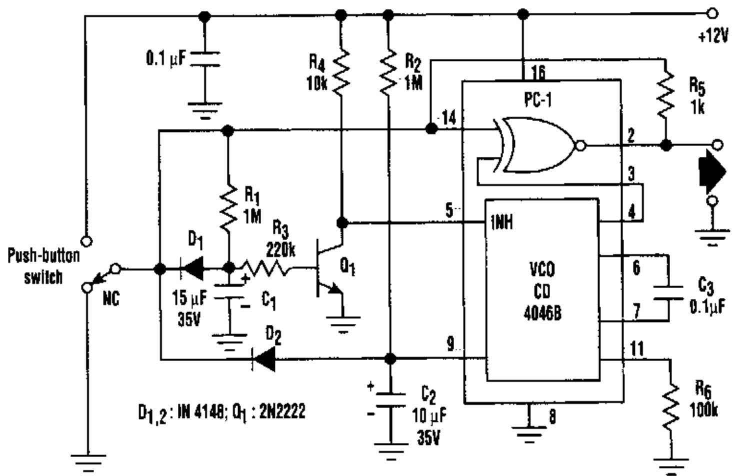
This staircase waveform first steps up and then steps down by the circuit shown. An input pulse generator provides the pulses that cause the output to step up or down, depending on the conduction of the clamp transistor, Q1. When this is ON, the "down" current pulse is diverted to ground and the staircase then steps "up." When the upper voltage trip point of Amp 2 is reached, Q1 goes OFF and as a result of the smaller "down" input resistor (one-half the value of the "up" resistor, R1), the staircase steps "down" to the low-voltage trip point of Amp 2. The output voltage, therefore, steps up and down between the trip voltages of the Schmitt Trigger.

87

Stepper Motor Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

STEPPER MOTOR PULSE GENERATOR



ELECTRONIC DESIGN

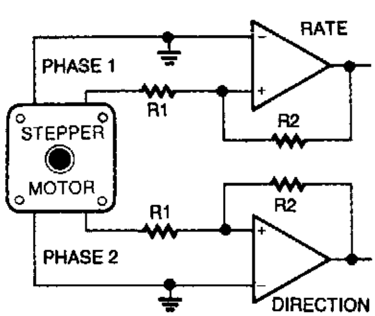
Fig. 87-1

When the switch is in its normally closed (NC) position, capacitors C1 and C2 are held discharged by diodes D1 and D2. Switching off transistor Q1 inhibits the voltage-controlled oscillator of the PLL. The two inputs and, hence, the output of the EX-OR gate (phase comparator 1) of the PLL remain at the logic 0 level.

When the pushbutton is pressed, C1 and C2 are allowed to charge via resistors R1 and R2. The VCO is enabled only after a time delay (≈0.5 second) set by R1, R3, and C1. During this delay period, the EX-OR gate output follows the logic level at the switch output. As a result, one-shot pulses can be generated by pressing the pushbutton, then releasing it within 0.5 second. R5 provides the switch-debouncing function.

If the pushbutton is pressed for more than 0.5 second, the VCO is enabled. The rising voltage at the control input (pin 9) causes a linear increase in VCO frequency and thus accelerates the stepper motor. Releasing the pushbutton discharges C1 and C2 and inhibits the VCO.

STEPPER MOTOR AS SHAFT ENCODER

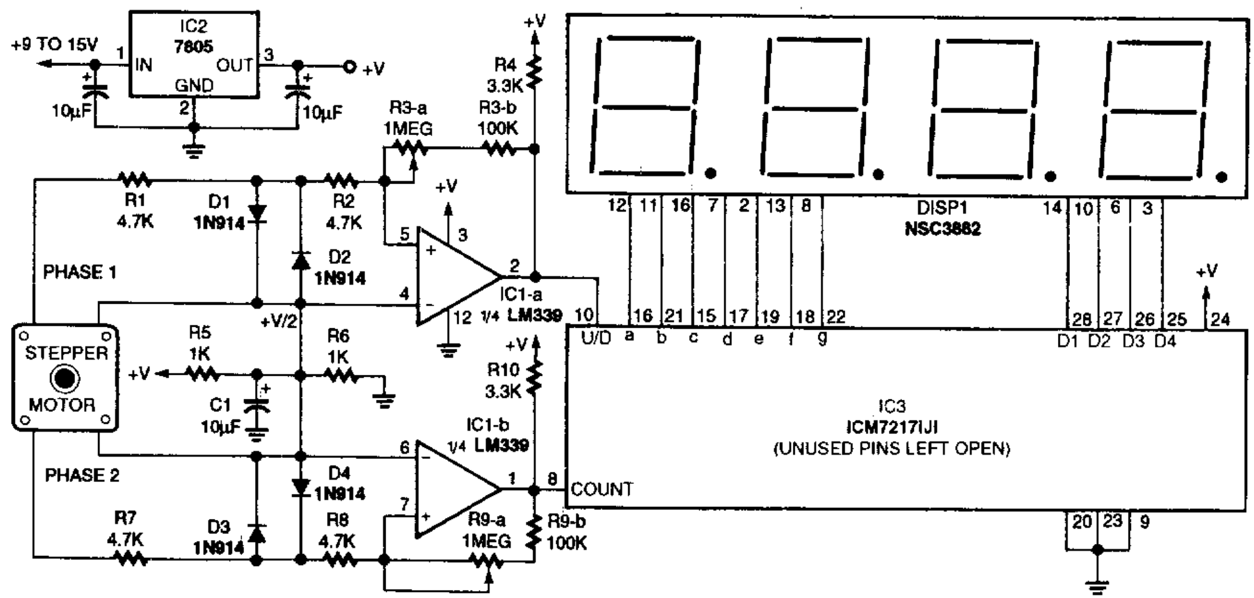


To use a stepper as a shaft encoder, the output signals must be converted to square waves with a pair of voltage comparators.

ELECTRONICS NOW

Fig. 87-2

STEPPER MOTOR ENCODER CIRCUIT



ELECTRONICS NOW

Fig. 87-3

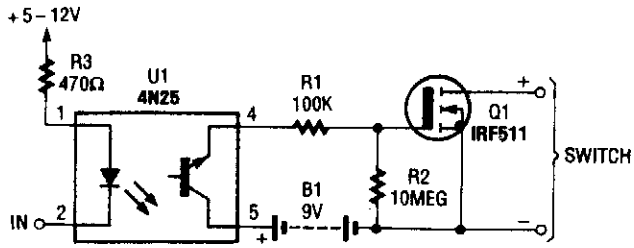
This circuit translates shaft rotation and direction to a readout on an LED display. A stepper motor is used as an encoder.

88

Switching Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

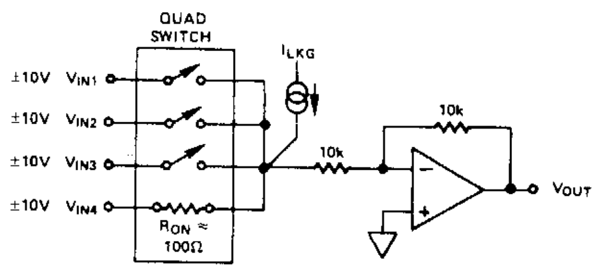
ISOLATED SWITCH



POPULAR ELECTRONICS ***Fig. 88-1***

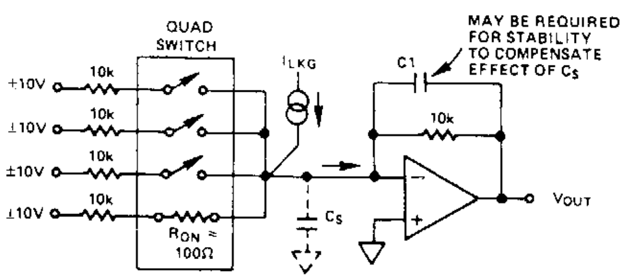
This dc-controlled switch uses an optoisolator/coupler, U1, to electrically isolate the input signal from the output control device.

ANALOG SWITCHED INVERTER



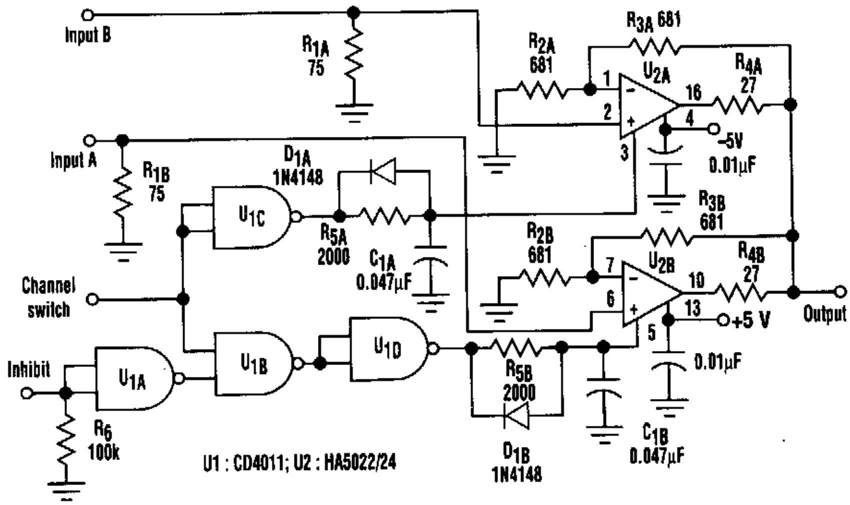
ANALOG DEVICES ***Fig. 88-2***

ANALOG SWITCH CIRCUIT



ANALOG DEVICES ***Fig. 88-3***

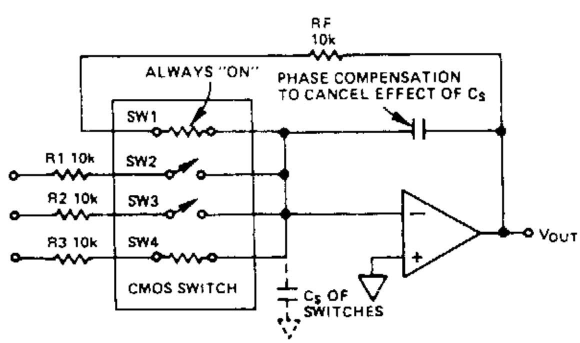
LOW OUTPUT IMPEDANCE MULTIPLEXER



ELECTRONIC DESIGN *Fig. 88-4*

Both inputs are terminated in their characteristic impedance; 75 Ω is typical for video applications. Because the output cables usually are terminated in their characteristic impedance, the gain is 0.5. Consequently, amplifiers U2A and U2B are configured in a gain of +2 to set the circuit gain at 1. R_2 and R_3 determine the amplifier gain; if a different gain is desired, R_2 should be changed according to the equation $G = (1 + R_3/R_2)$. R5, LCL1, and D1 make up an asymmetrical charge/discharge time circuit that configures U1 as a break-before-make switch to prevent both amplifiers from being active simultaneously. The multiplexer transition time is approximately 15 μs with the component values shown.

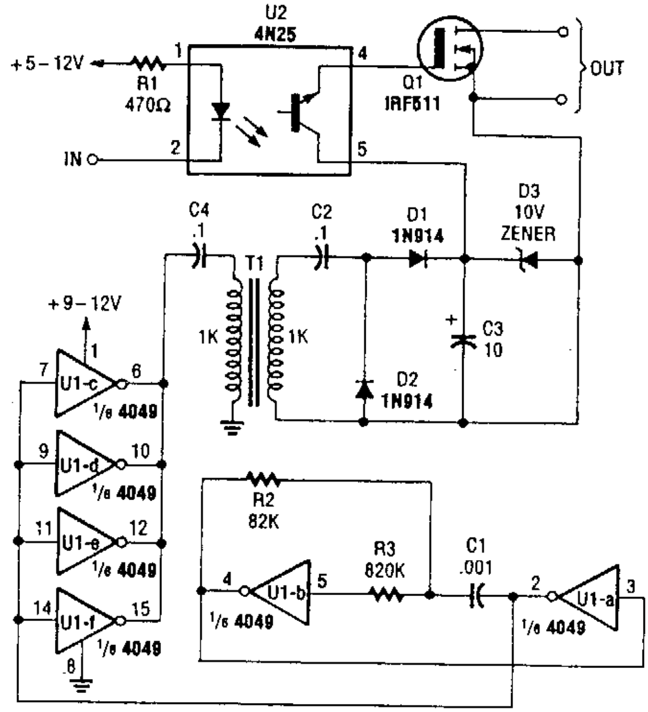
OP-AMP AND ANALOG SWITCH RON COMPENSATOR



ANALOG DEVICES *Fig. 88-5*

This switch is in series with feedback resistor to compensate gain.

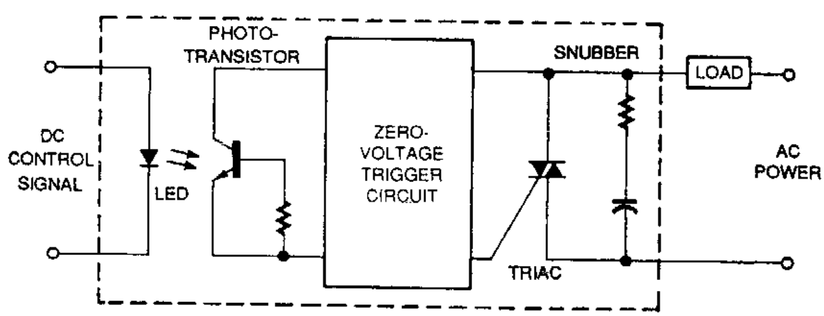
OSCILLATOR TRIGGERED SWITCH



POPULAR ELECTRONICS *Fig. 88-6*

In this circuit (the oscillator-triggered switch), the HEXFET's base bias is provided by a signal generated by an astable oscillator.

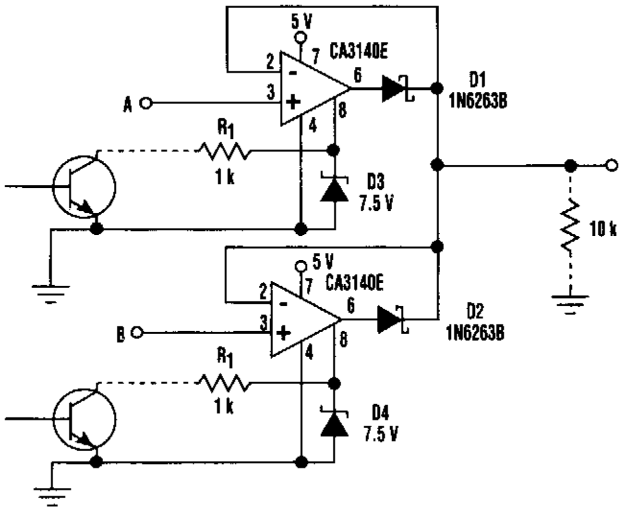
BASIC ZERO-CROSSING SWITCH CIRCUIT



ELECTRONICS NOW *Fig. 88-7*

Here is the schematic of a solid-state ac relay with zero-crossing. The triac permits the relay to switch to ac directly.

ANALOG SWITCH



ELECTRONIC DESIGN

Fig. 88-8

This design takes advantage of the strobed output stage of a CA3140 amplifier. With the strobing capability, the circuit's output voltage can be set to either of the input voltages by grounding one of the control inputs, either A or B.

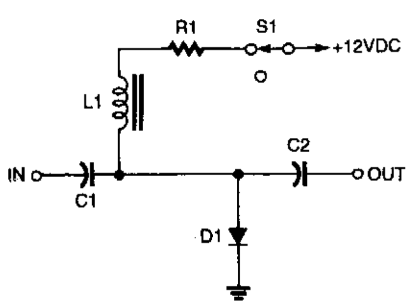
When the strobe input at pin 8 is taken below 1 V, that amplifier is disabled. The remaining amplifier then acts as a unity-gain high-impedance buffer.

The 10-kΩ output resistor enables the output voltage to swing down to 20 mV from ground. The Zener-diode clamps and associated resistors connected at the strobe inputs can be omitted for the lowest-cost applications. However, experience has shown that they allow the amplifiers to shrug off the effect of high transient voltages.

The circuit is particularly suited to 8-bit microcontroller applications, where the strobe inputs can be driven directly from two open-collector output ports under software control.

The use of Schottky diodes for D1 and D2 makes possible an output swing of 2.5 V when the circuit is powered from a 5-V supply.

SHUNT PIN-DIODE SWITCH

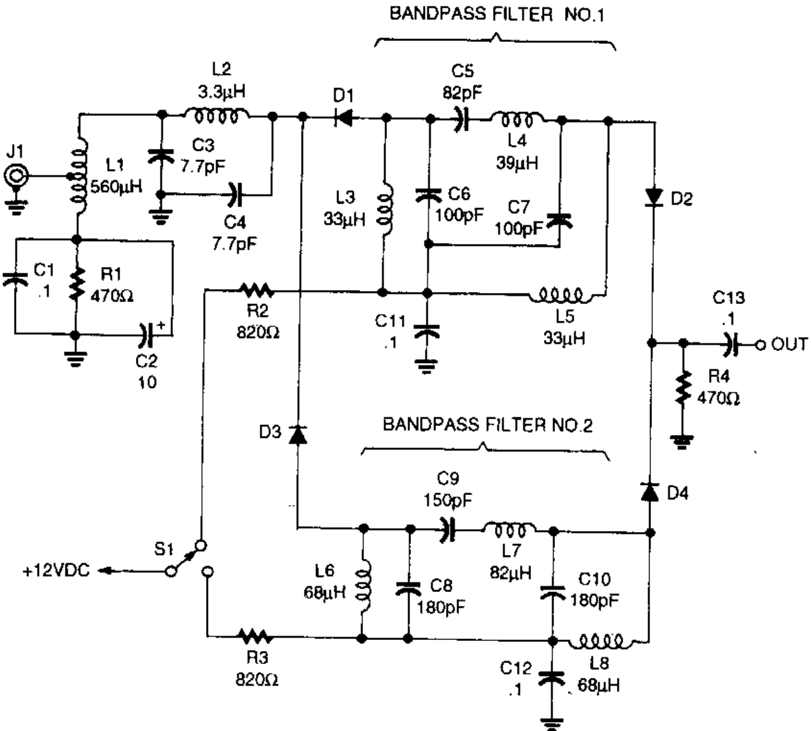


This PIN-diode switching circuit directs signals to ground when D1 is forward-biased. R1 is typically 470 Ω to 2.2 kΩ. $C_1 = C_2 = 0.1$ μF.

POPULAR ELECTRONICS

Fig. 88-9

RECEIVER BANDSWITCHING

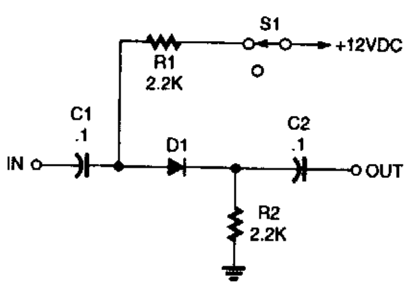


POPULAR ELECTRONICS

Fig. 88-10

Eight-band receiver front-end selection can be accomplished by using PIN diode switches.

RESISTOR PIN-DIODE SWITCH

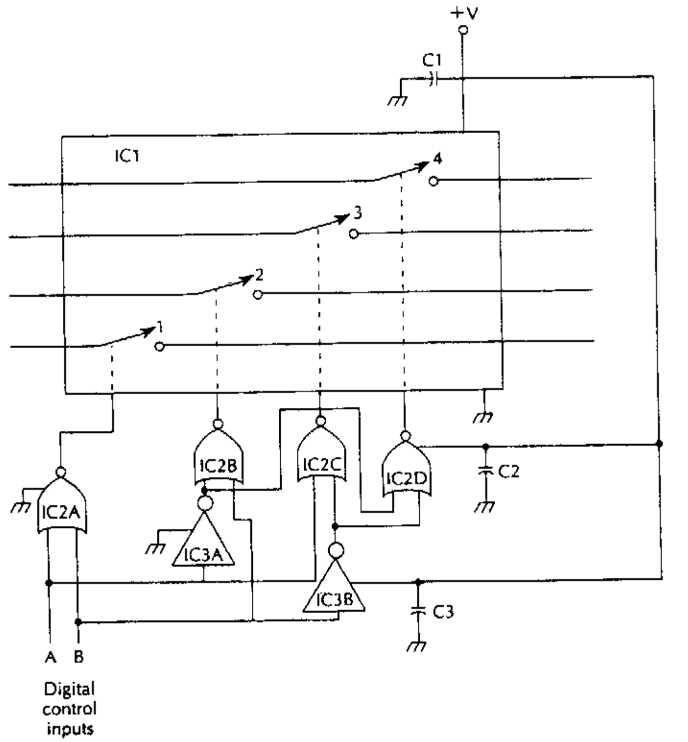


POPULAR ELECTRONICS

Fig. 88-11

This circuit uses resistors instead of RF chokes to keep costs low. The values of R_1 and R_2 should be no lower than about 1 kΩ to minimize loss.

DIGITALLY CONTROLLED ONE-OF-FOUR ANALOG SWITCH

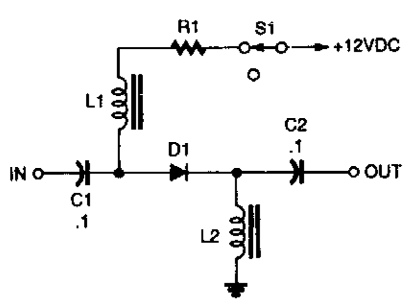


McGRAW-HILL

Fig. 88-12

IC1 = CD4066
IC2 = CD4001
IC3 = CD4049
All capacitors = 0.1 μF

PIN DIODE SWITCH

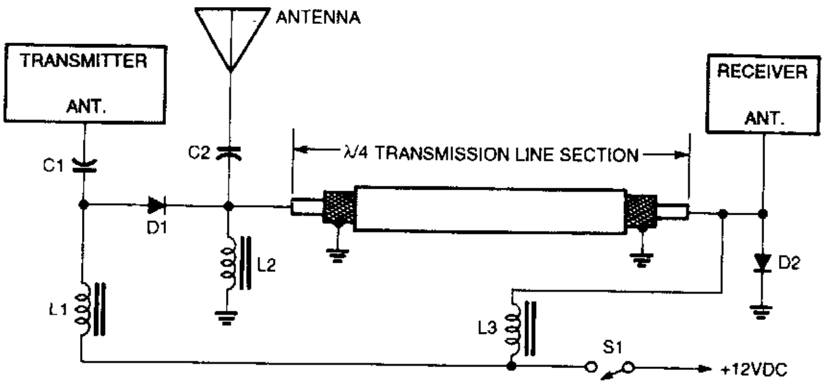


POPULAR ELECTRONICS

Fig. 88-13

This PIN diode switch uses RF chokes and a single diode. R1 is typically 470 Ω to 2.2 kΩ.

TRANSCEIVER T/R SWITCH

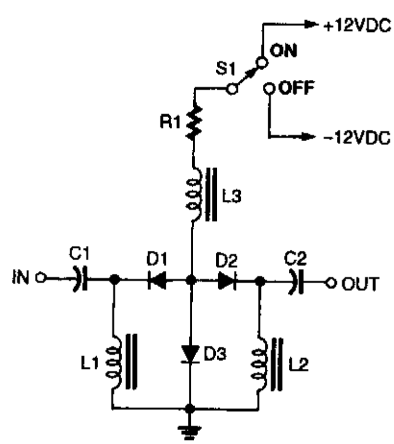


POPULAR ELECTRONICS

Fig. 88-14

This transceiver's transmit/receive switch uses PIN-diode instead of a relay. On receive, D1 is cut off, D2 is cut off and the antenna connects to the receiver. During transmit, D1 is forward-biased, as is D2. This connects the receiver input. This causes the input impedance of the transmission line to be high, so little transmitter power reaches the receiver. Although not shown in the schematic, the 12-V supply should have a series resistor of 100 Ω to 2.2 kΩ, depending on diode current, to limit diode current to a safe value.

SERIES/SHUNT PIN-DIODE RF SWITCH

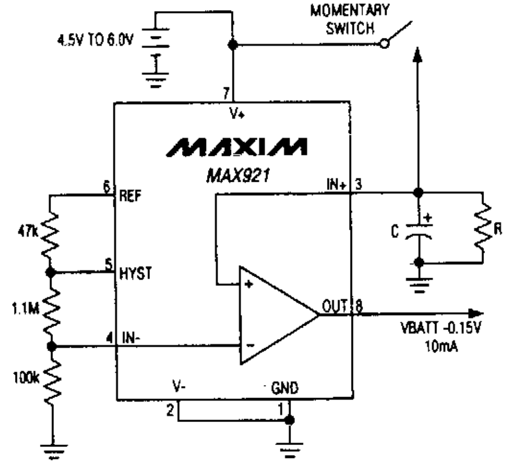


POPULAR ELECTRONICS

Fig. 88-15

A combination of series and shunt switching, like that shown here, results in superior isolation between the input and output when in the off condition.

AUTO-OFF POWER SWITCH

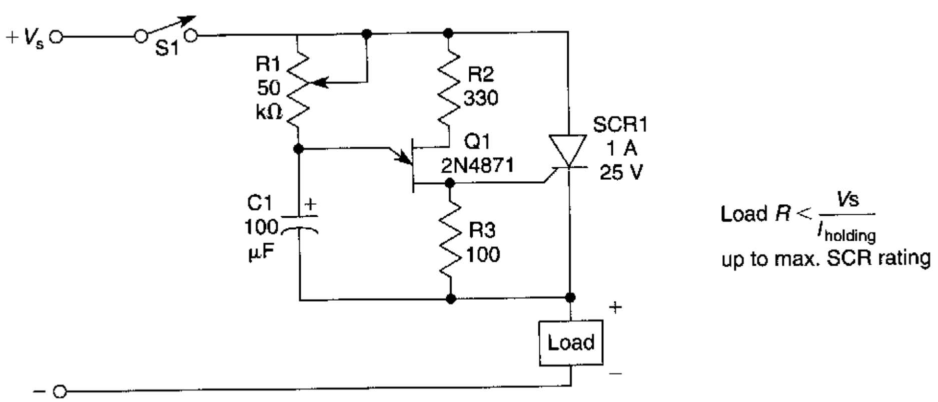


MAXIM

Fig. 88-16

This timed switch circuit can be used where a timed power source is needed. The on-time is approximately 4.6 *RC*.

SWITCH-ON DELAY CIRCUIT

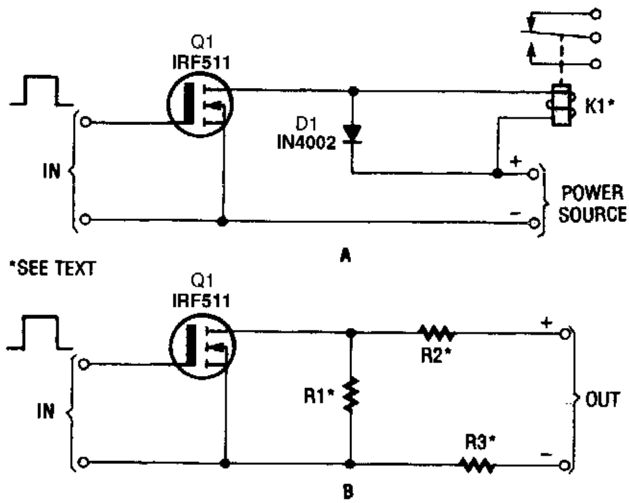


WILLIAM SHEETS

Fig. 88-17

When S1 is turned on, a very small current flows through the load. Almost the entire supply voltage appears across the SCR. When C1 changes up to the firing voltage of Q1 (approximately the standoff ratio of Q1 times V_s usually 0.4 to 0.6 V_s) through R1, Q1 fires, turning on SCR1. This delivers full voltage to load, minus SCR drop (about 1.2 V). Notice that load current must exceed SCR holding current.

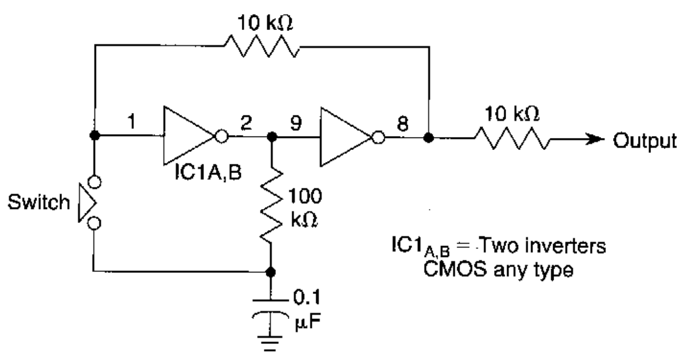
HEXFET SWITCH CIRCUITS



POPULAR ELECTRONICS *Fig. 88-18*

The HEXFET can switch dc power to relays (as shown in A), motors, lamps, and numerous other devices. That arrangement can even be used to switch resistors in and out of a circuit, as shown in Fig. 88-18B. R1, R2, and R3 are possible load resistors and represent load configurations that can be used.

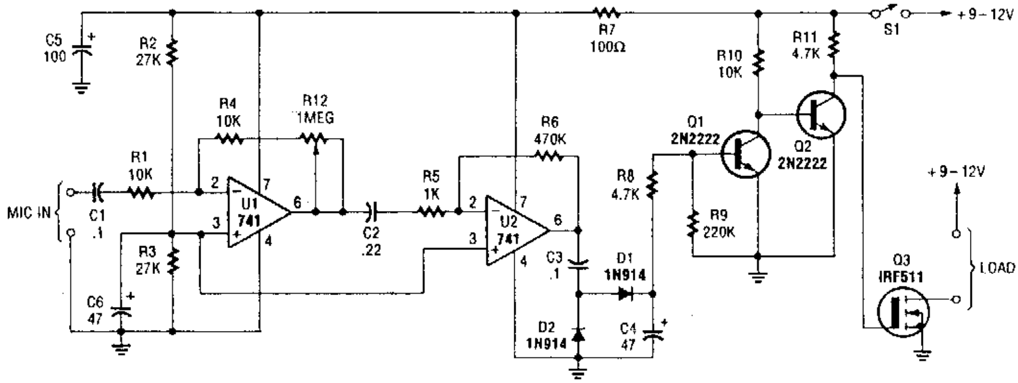
ALTERNATING ON/OFF CONTROL



WILLIAM SHEETS *Fig. 88-19*

When the switch is closed, it causes a change in the state of pins 1 through 8. This will provide a toggle flip-flop action.

AUDIO-CONTROLLED SWITCH

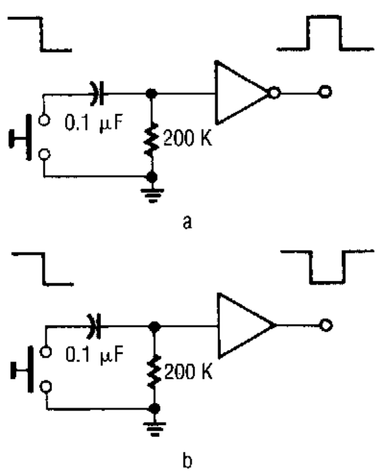


POPULAR ELECTRONICS

Fig. 88-20

The audio-controlled switch combines a pair of 741 op amps, two 2N222 general-purpose transistors, a HEXFET, and a few support components to produce a circuit that can be used to turn on a tape recorder, a transmitter, or just about anything using sound.

SWITCH DEBOUNCERS

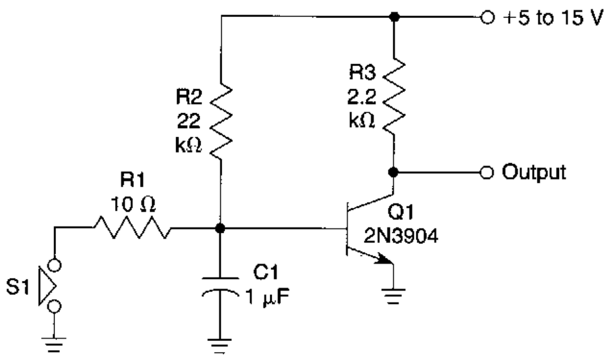


These circuits will cure problems caused by switch-contact bounce. The one shown in Fig. 88-21A provides you a positive output pulse, and the one shown in Fig. 88-21B provides you a negative output pulse.

ELECTRONICS NOW!

Fig. 88-21

SIMPLE SWITCH DEBOUNCER

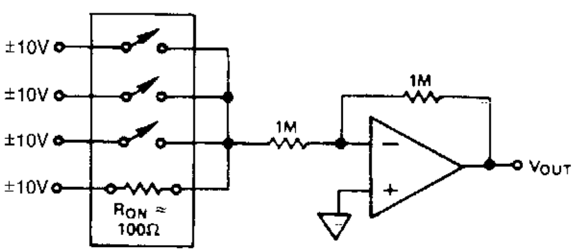


Pressing S1 discharges C1 through R1, causing Q1 to cut off, forcing the output high. Once C1 is discharged below the V_{BE} (ON) of Q1, switch bounce will have no effect on the output.

WILLIAM SHEETS

Fig. 88-22

ANALOG SWITCH CIRCUIT



ANALOG DEVICES

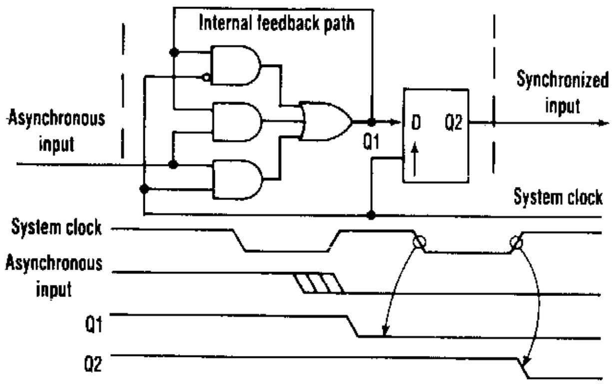
Fig. 88-23

89

Sync Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

PLD SYNCHRONIZES ASYNCHRONOUS INPUTS



ELECTRONIC DESIGN *Fig. 89-1*

A programmable electrically erasable logic (PEEL) device can easily supply the synchronizing function. Digital systems often require synchronization of asynchronous inputs to avoid the potential metastability problems caused by setup-time violations. A common synchronization method uses two rippled 74LS72 D-type flip-flops.

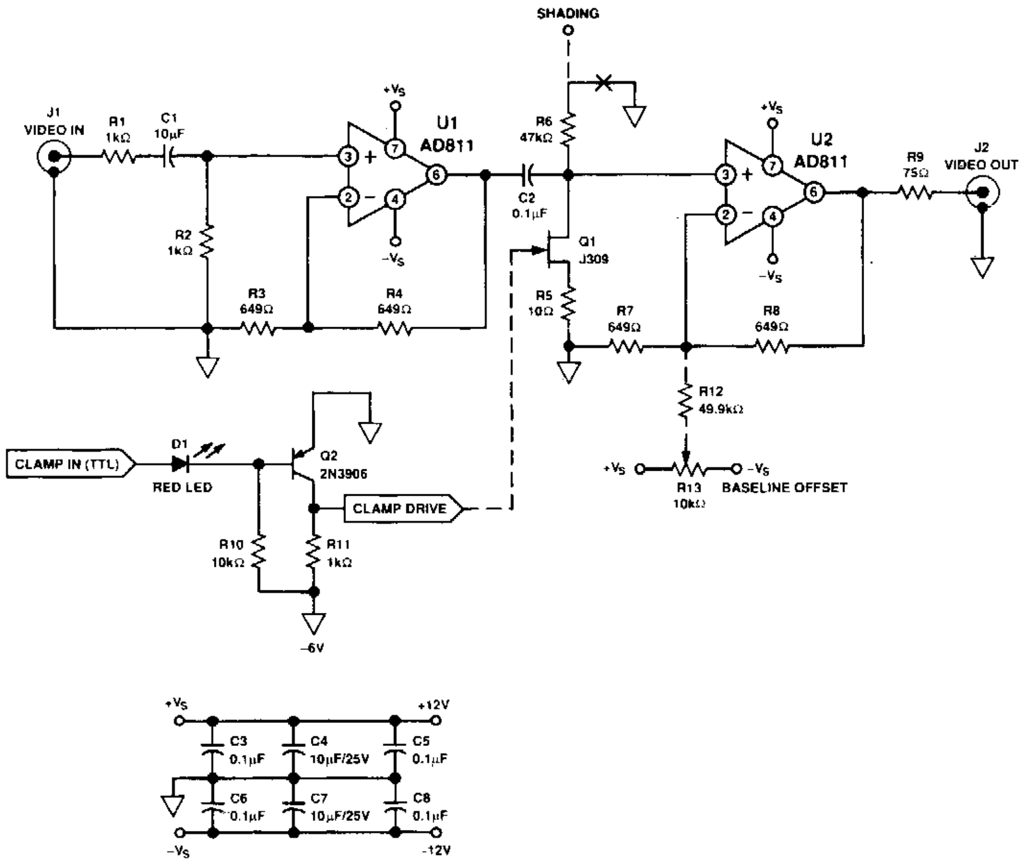
In this circuit, the asynchronous input feeds into the D input of the first flip-flop and its Q output feeds into the D of the second. Because the first flip-flop latches on the falling edge of the system clock, to avoid setup-time violations, the D input signal to the second flip-flop will be stabilized before the rising edge of the clock. Even experienced programmable-logic device designers often resort to such a TTL flip-flop circuit to handle the synchronization function, because of the architectural limitations of standard PLDs.

A programmable electrically erasable logic (PEEL) device, such as the PEEL18CV8 from ICT, however, can easily supply the function. The user-programmable 12-configuration I/O macrocells in the device can internally feed back a signal before the output register. With this feedback arrangement, designing a two-stage input is simple.

A gated-latch internally latches the asynchronous input on the falling edge of the system clock, generating signal Q1. ANDing the input with Q1 through the internal feedback path, eliminates a possible hazard condition during the clock's high-to-low transition time. The latch then holds Q1 stable to ensure meeting the setup-time requirement of the subsequent D flip-flop, which, as before, registers the signal on the next rising system clock edge.

If by chance the input pulse width violates the set-up time of the gated latch, the clock's low time will give more time for settling.

SYNC TIP dc RESTORER

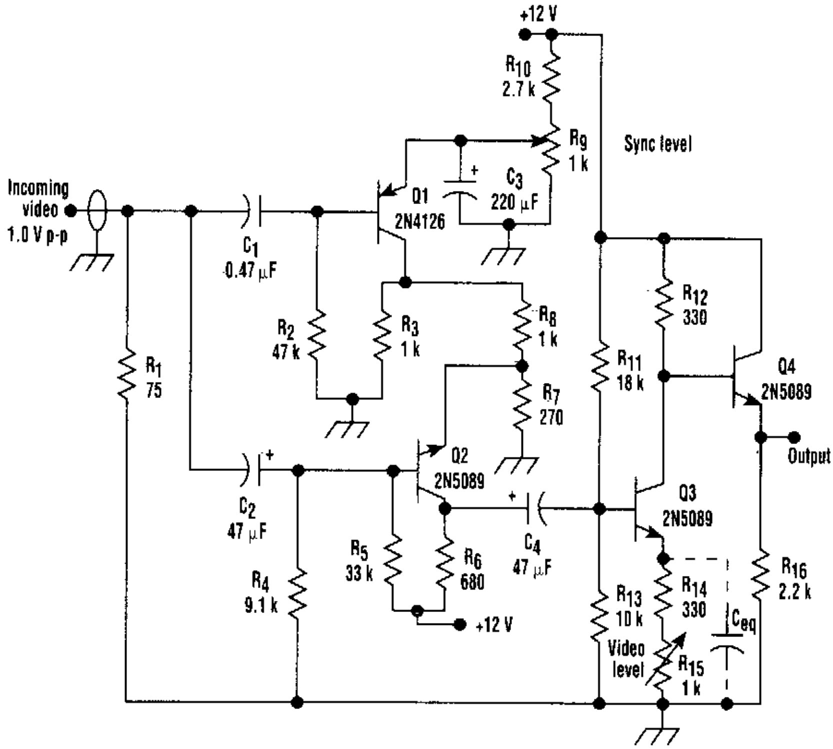


ANALOG DEVICES

Fig. 89-2

The dc restorer shown supplies a video signal with sync tips clamped to a baseline level. Clamp drive signal is supplied from elsewhere, usually a sync generator or a sync separator.

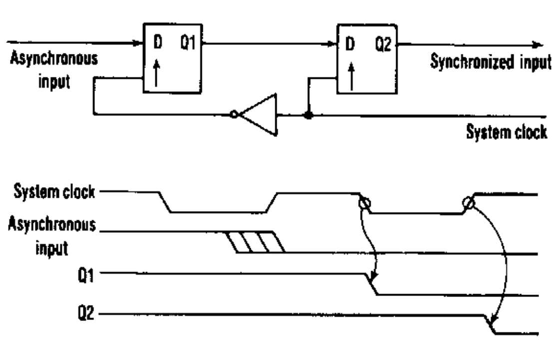
SYNC STRETCHER CIRCUIT



ELECTRONIC DESIGN *Fig. 89-3*

Q1, Q2, and Q3 comprise a simple video amplifier and sync stretch circuit. Transistor Q1 sync strips the incoming video, which is amplified and mixed with the stripped sync in Q2. Q3 supplies inversion and video amplitude control.

SYNCHRONIZER CIRCUIT



ELECTRONIC DESIGN *Fig. 89-4*

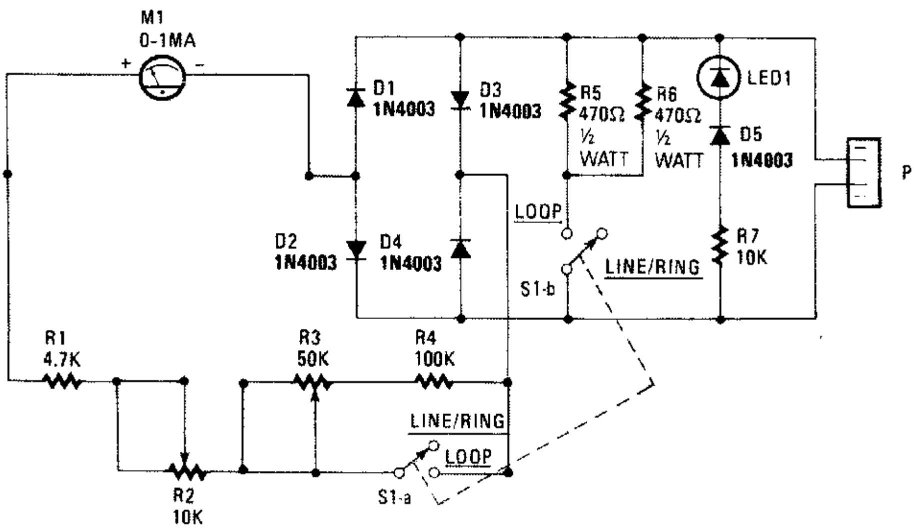
This common synchronization method uses two rippled 74LS74 D-type flip-flops.

90

Telephone-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

TELEPHONE LINE TESTER



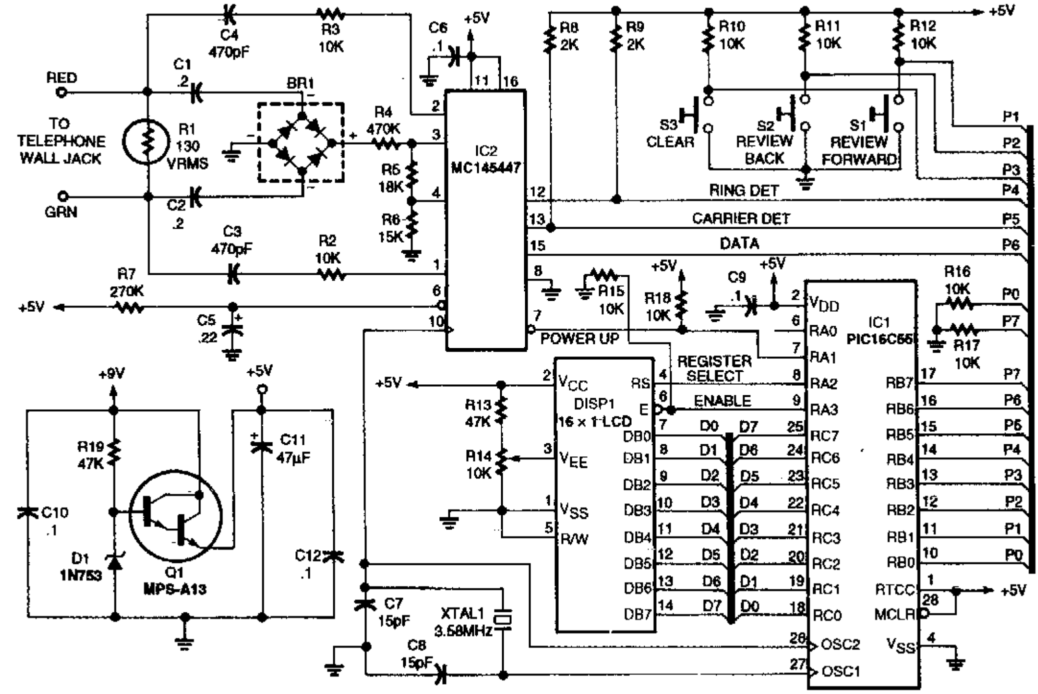
RADIO-ELECTRONICS *Fig. 90-1*

The telephone-line tester shown in the figure is connected to the telephone line through modular connector P1. Because the tester's LED polarity indicator is always connected when the tester is plugged in, the instant that the unit is connected, you will have an indication of the polarity. If it is correct—that is, if the green wire is the positive side and the red wire is the negative side—nothing will happen. If the situation is reversed, the LED will light.

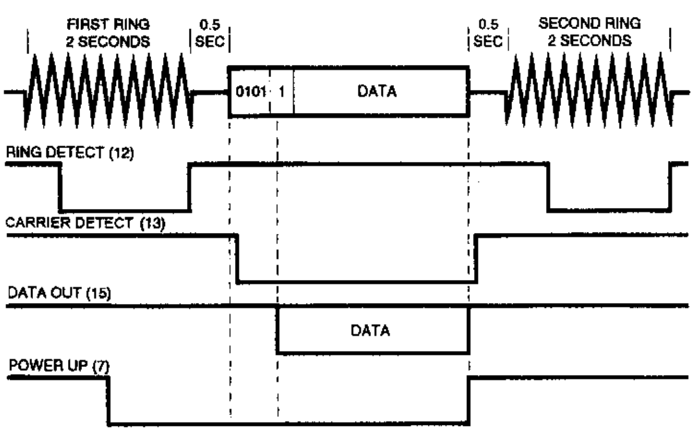
With switch S1 set for LINE/RING, both S1-a and S1-b are open and the meter indicates the condition of the line voltage. Any line-voltage reading in the LINE OK range (more on the meter in a moment) indicates a line voltage that is higher than 40 Vdc. If the telephone is caused to ring, either by using a ringback number or by dialing from another phone, the meter will indicate RING OK, and the LED will pulse (indicating ac), if the ringing voltage/current is correct. The actual position of the meter's pointer depends on how many ringers are connected across the line.

When S1 is closed the voltage range of the meter is changed and a nominal load resistance of 230 Ω (R5 and R6) is connected across the line to emulate the off-hook load of the telephone. If the meter indicates LOOP OK, you can be certain that you have sufficient loop voltage for satisfactory telephone operation. If you place another load on the line, perhaps by taking an extension telephone off hook, the meter reading will almost invariably drop below the LOOP OK range. If lifting the handset causes the meter reading to drop, you can at least be certain that the telephone's hook switch is working and that the repeat coil is connected to the line.

CALLER ID CIRCUIT



THE HEART OF THE CALLER ID circuit is microcontroller IC1 which processes the serial data from IC2, outputs ASCII characters to DISP1, and monitors switches S1–S3.

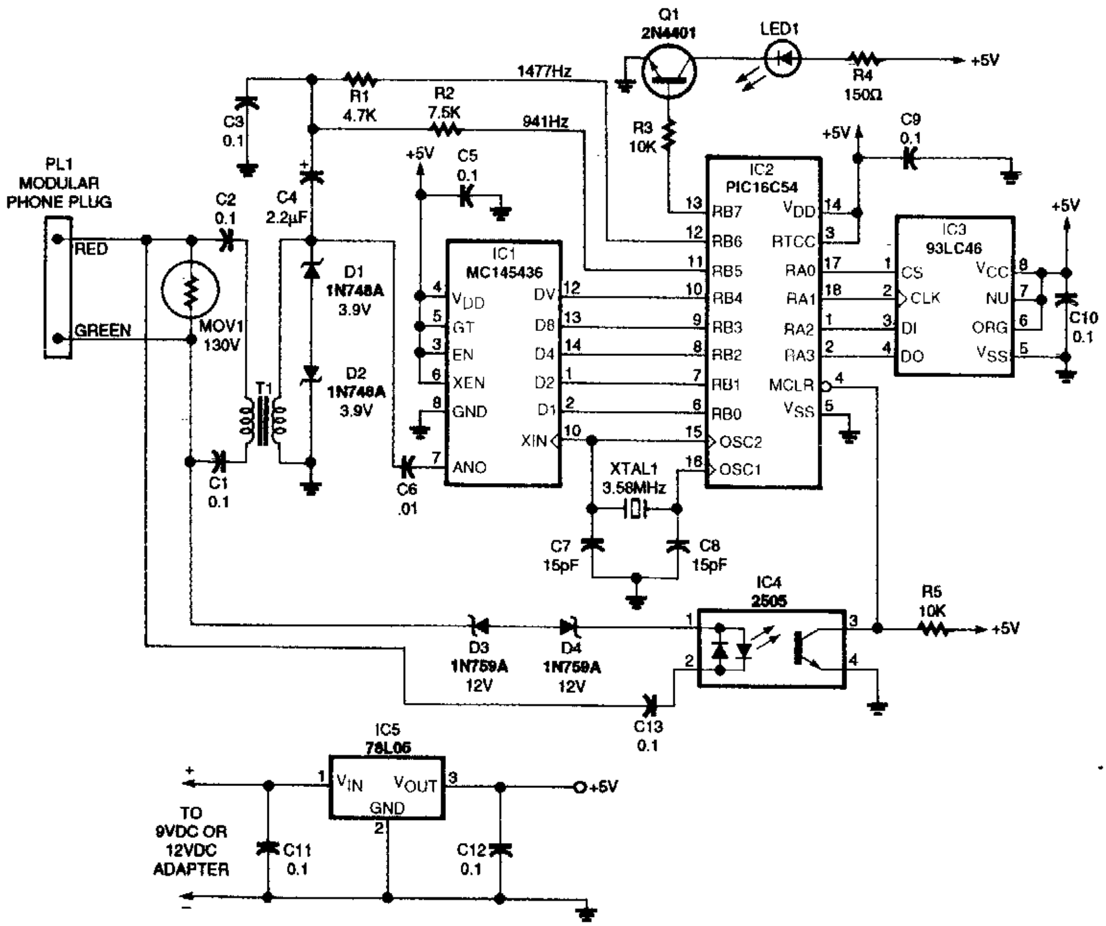


TIMING RELATIONSHIPS between the data present on the phone line (top), and the output pins of IC2.

ELECTRONICS NOW *Fig. 90-2*

This circuit requires programming of the microcontroller. Software information is available from the reference in the original article.

TELEPHONE CALL RESTRICTOR

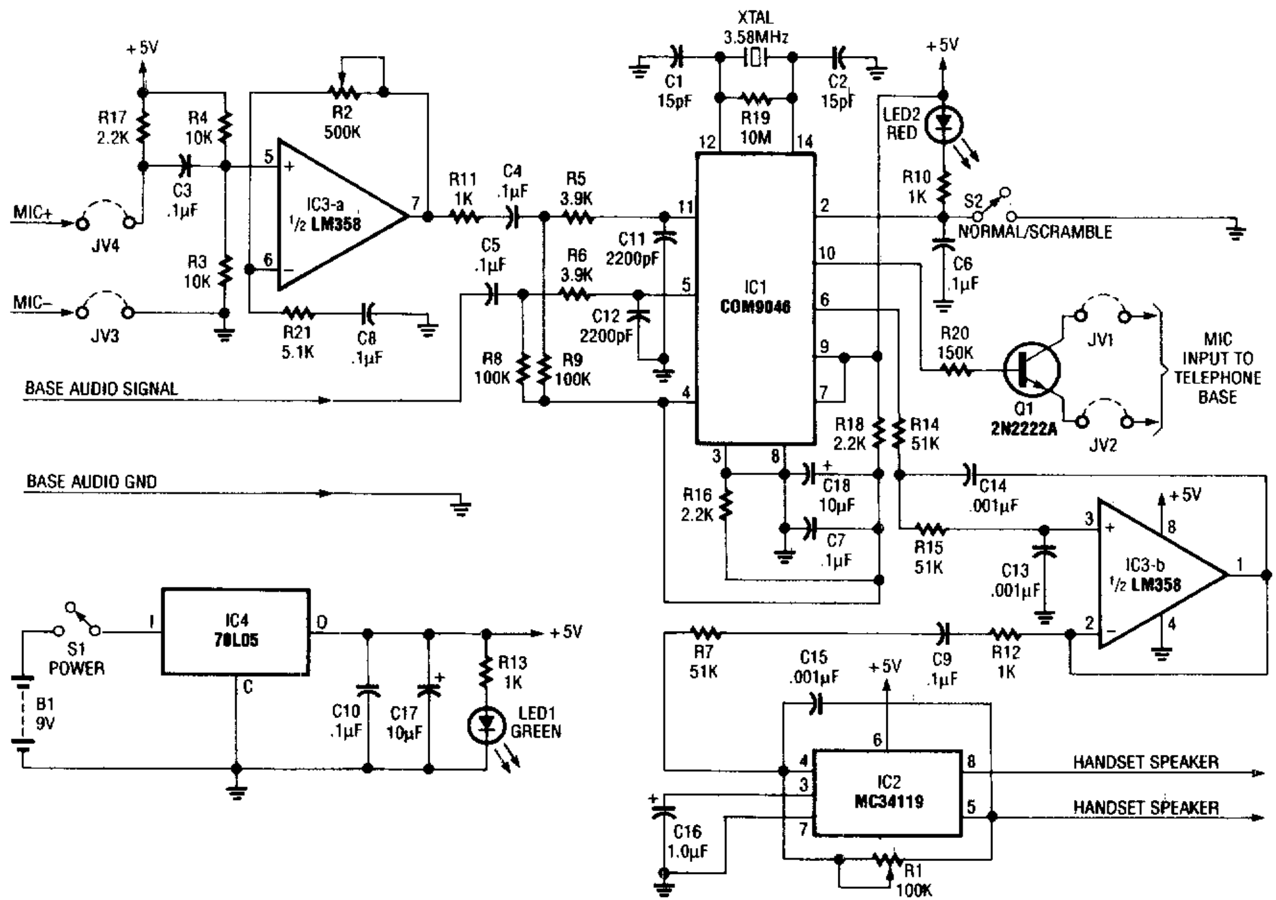


ELECTRONICS NOW

Fig. 90-3

This circuit is designed to restrict phone calls with the area codes: 900, 976, and 540. This device uses a microcontroller to compare the DTMF decoded tones with telephone numbers stored in EEPROM (IC3). This device requires a programmed microcontroller. Software and details of programming can be found in the original magazine article.

TELEPHONE SCRAMBLER

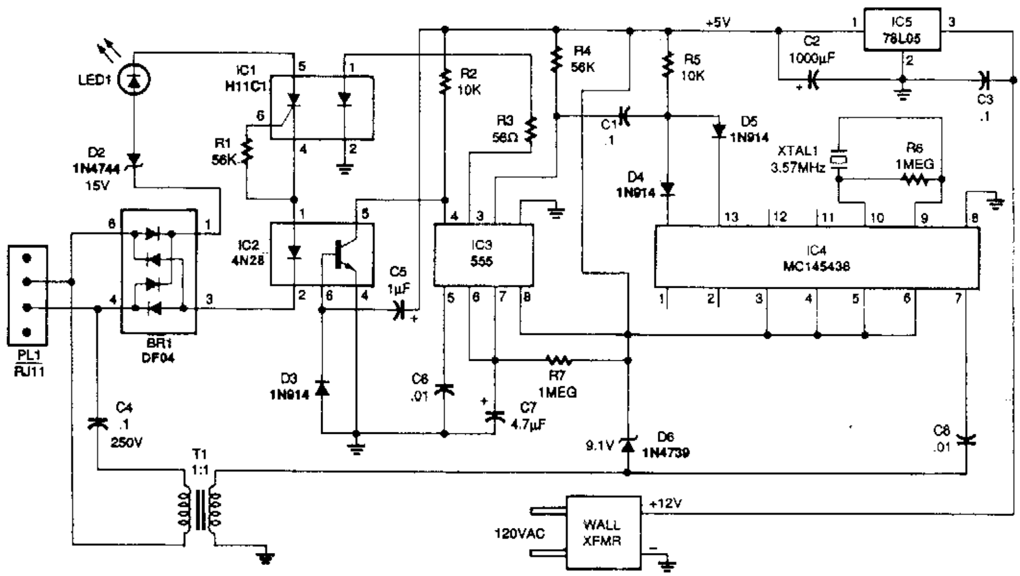


ELECTRONICS EXPERIMENTERS HANDBOOK

Fig. 90-4

This circuit uses the usual speech inversion algorithm, implementing it with a COM9046 ASIC. This unit is designed to fit between the handset and base of a standard telephone. It is powered by a 9-V battery.

UNIVERSAL TELEPHONE HOLD CIRCUIT

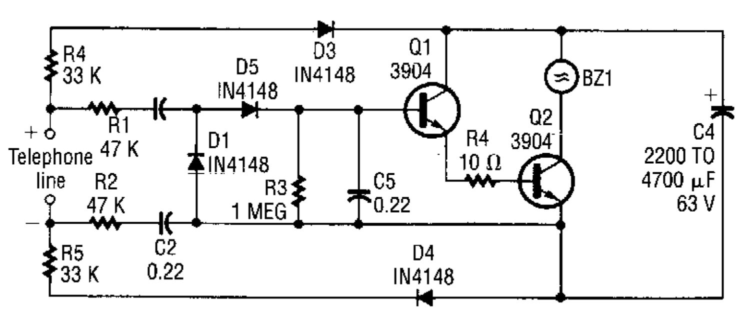


ELECTRONICS NOW

Fig. 90-5

The telephone line is connected to the hold components through bridge rectifier BR1 so that the input is not polarity sensitive. If you have touch-tone telephone service, you can now put a call on hold from any phone in your house by plugging this simple device into any telephone jack. The universal hold-circuit works with any phone that has a key pad with a # key. To put a call on hold, press the # key and hang the phone up. A timer extends the #-key function while you hang up phones that have a keypad built into the handset.

TELEPHONE RING AMPLIFIER

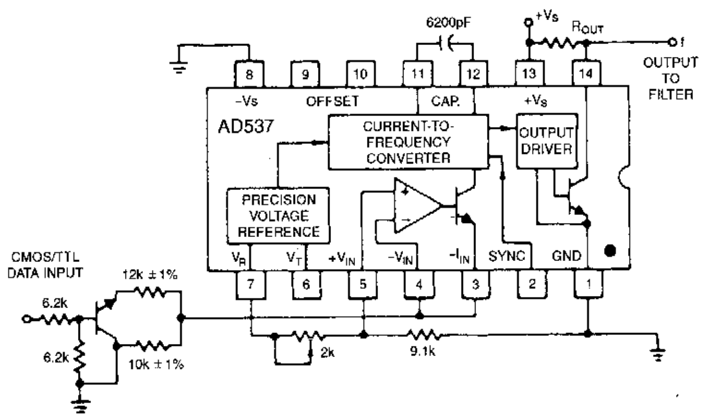


ELECTRONICS NOW

Fig. 90-6

This circuit takes its operating power from the telephone line. BZ1 is a piezoelectric transducer.

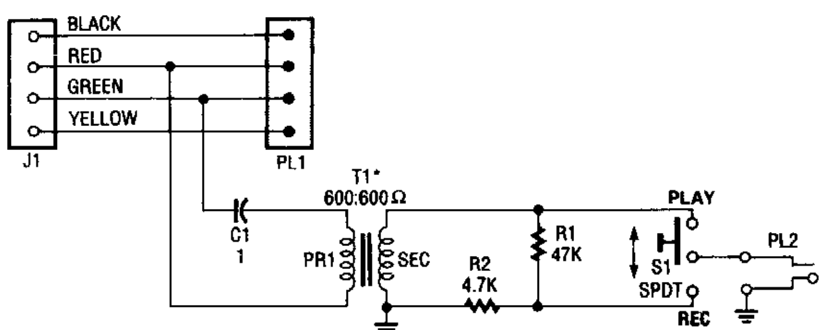
BELL SYSTEM 202 DATA ENCODER



ANALOG DEVICES *Fig. 90-7*

The AD537 is well-suited for frequency-shift modulator and demodulator applications. Requiring little power, it is especially appropriate for using phone-line power. The Bell-System 202 data encoder shown here delivers the mark frequency of 1.2 kHz with the data input low. When the input goes high, the timing current increases to 165 μA and generates the space frequency of 2.2 kHz. The trim shown provides a ±10% range of frequency adjustment. The output goes to the required bandpass filter before transmission over a public telephone line. A complementary demodulator is easy to implement.

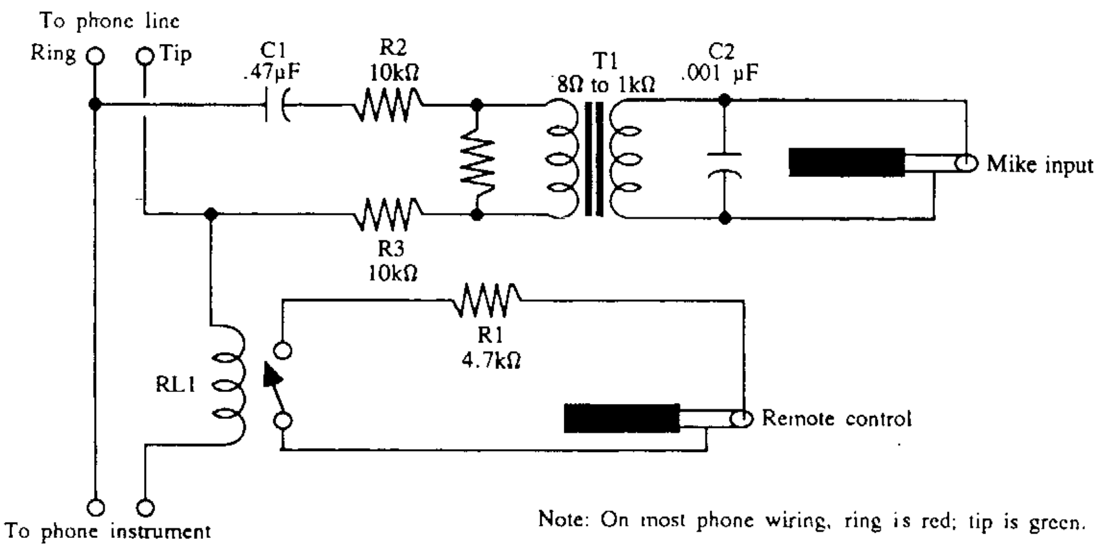
TELEPHONE/AUDIO INTERFACE



POPULAR ELECTRONICS *Fig. 90-8*

The telephone audio interface—essentially, a simple isolation/couple circuit—isolates the phone line from any connected audio circuit without presenting any danger to the phone line, the equipment, or the user.

TELEPHONE RECORDING CIRCUIT



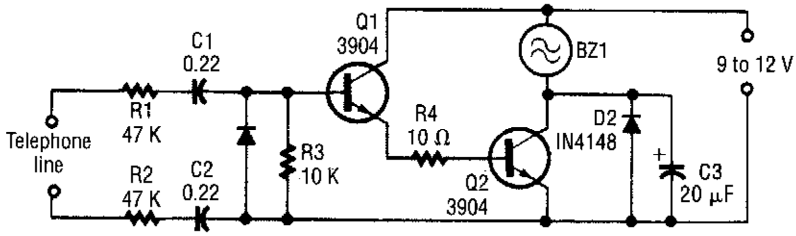
R1	4.7 kΩ
R2,R3	10 kΩ
C1	0.47 μF disc
C2	0.001 μF disc
T1	8 kΩ-to-1 kΩ impedance-matching transformer
RL1	SPST reed relay

All resistors are 5 to 10 percent tolerance, 1/4 watt. All capacitors are 10 to 20 percent tolerance, rated at 35 volts or more.

McGRAW-HILL *Fig. 90-9*

This device will automatically record telephone calls. An ordinary cassette recorder can be hooked to it.

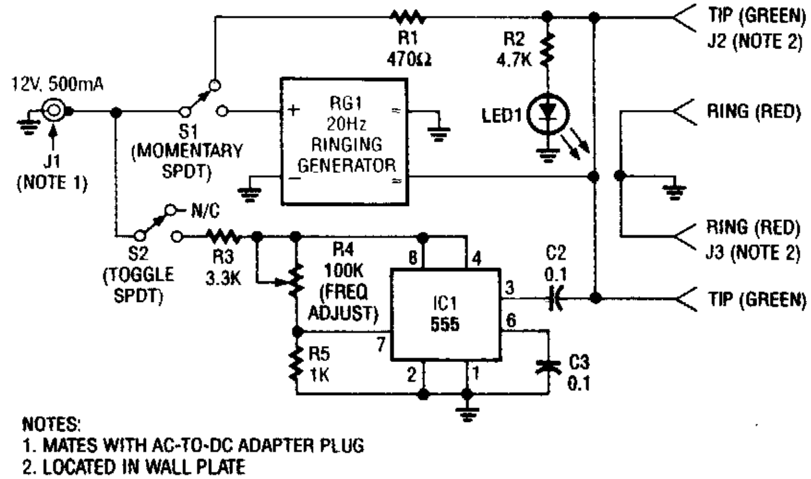
TELEPHONE BELL AMPLIFIER



ELECTRONICS NOW *Fig. 90-10*

Telephone "bell" amplifier circuit will let you hear (or see) an enhanced alarm if you are away from your telephone.

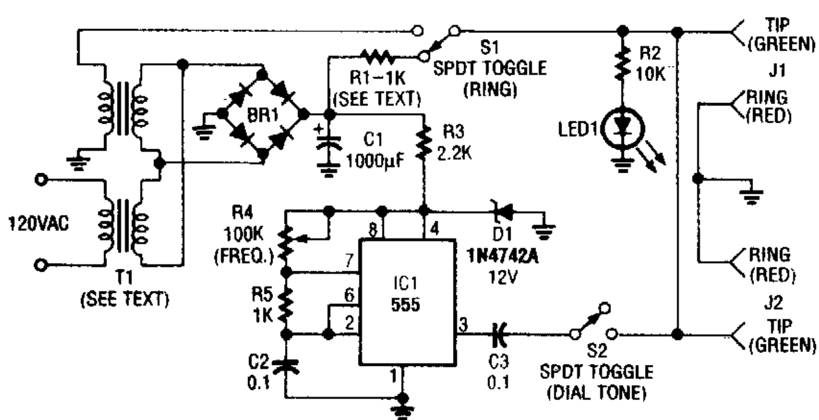
PHONE LINE SIMULATOR



ELECTRONICS NOW *Fig. 90-11*

This device contains a ringing generator and a tone oscillator. The tone oscillator is set to either 350 or 440 Hz. The ringing generator is a potted module delivering 86 Vac at 20 Hz. It is available from a source listed in the reference.

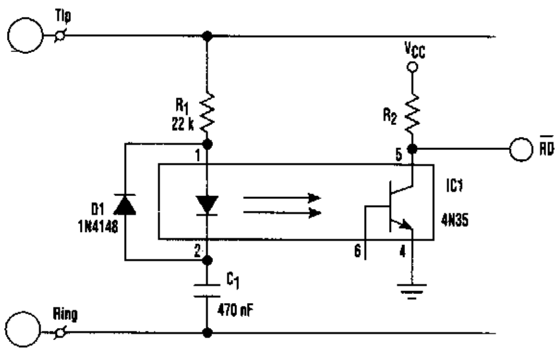
PHONE HELPER



RADIO-ELECTRONICS *Fig. 90-12*

This phone-line simulator uses a 60-Hz transformer instead of a ring generator. The dial tone is provided by an NE555 astable oscillator.

TELEPHONE RING SIGNAL DETECTOR



ELECTRONIC DESIGN *Fig. 90-13*

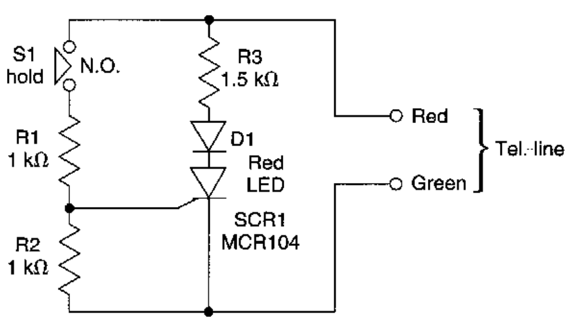
Discriminating between telephone-ring signals on a phone line can be accomplished by using dedicated ICs, such as AT&T's LB1006AB or Texas Instruments' TMS1520A. However, if the system already is using a microcontroller, those dedicated chips can be replaced with simpler hardware and a few bytes of code.

Looking at the setup, the ringing voltage pulses the optoisolator's LED; which, in turn, pulses the low-asserted RD line to the microcontroller. The firmware analyzes the pulses to determine whether a valid ringing signal is present.

The frequency limits of a valid signal are 20 to 80 Hz, which is modulated 2 seconds on and 4 seconds off (with distinctive ringing, though, this cadence can vary). Therefore, the simplest analysis is to count down at least 20 pulses of RD in 1 second.

The routine could be expanded to determine what type of ring signal is present in a distinctive ring setting. Such a system could switch the phone line to various output jacks. As a result, several phone devices could use the same line without first picking up the line to determine if it's a voice, fax, or data call.

TELEPHONE HOLD CIRCUIT



When the hold button S1 is pressed, SCR1 fires via R1 and R2, firing SCR1, and seizing the line via the path through R3, D1, and SCR1.

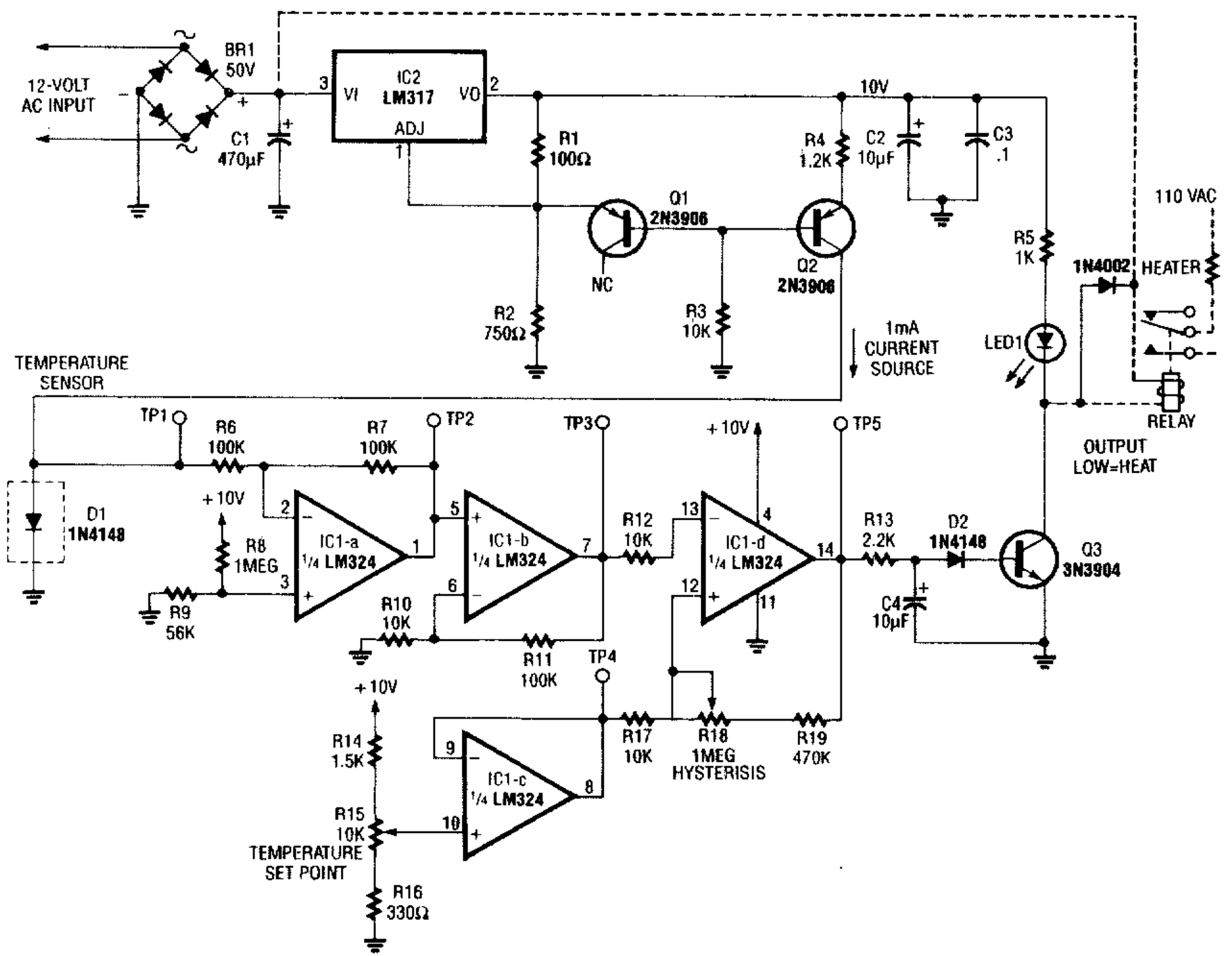
WILLIAM SHEETS *Fig. 90-14*

91

Temperature-Related Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ELECTRONIC THERMOSTAT

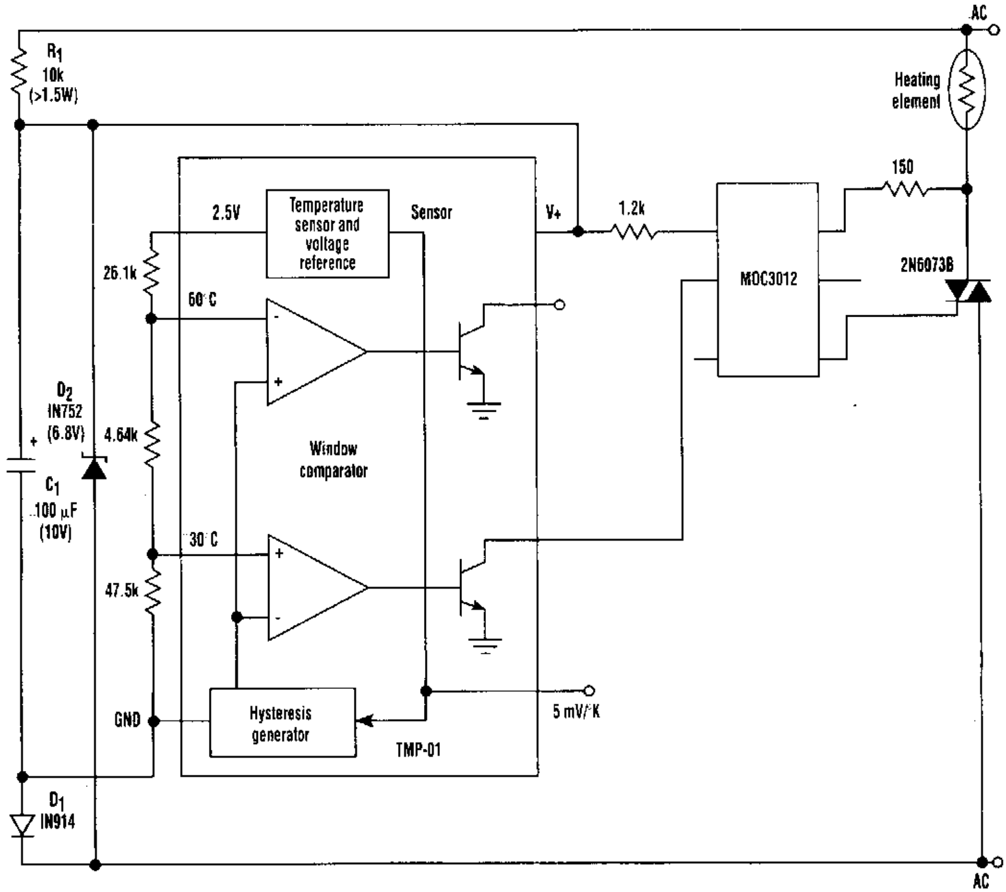


ELECTRONICS EXPERIMENTERS HANDBOOK

Fig. 91-1

Using a 1N914 diode as a temperature sensor, this straightforward circuit has hysteresis and set-point adjustments. Usable range is about –50 to +150°C.

TEMPERATURE CONTROLLER



ELECTRONIC DESIGN

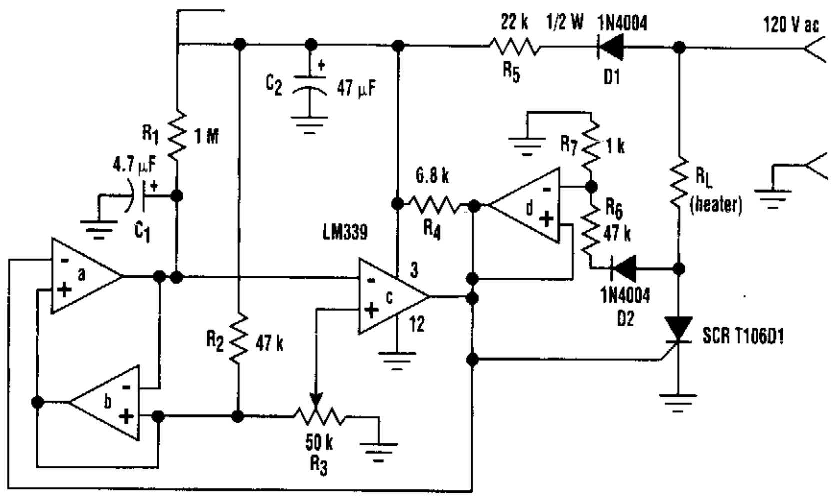
Fig. 91-2

The temperature sensor/controller (the TMP-01) is a monolithic device whose low power allows it to operate with a simple half-wave rectified power supply directly from the ac line. Such an arrangement greatly simplifies the power-supply design requirement to the point of only needing a few low-cost components to provide a single +6-Vdc supply.

The TMP-01 is essentially a "thermostat on a chip." It includes a linear temperature sensor (5 mV/K), and also has two comparators that switch at externally determined set points. These set points are established by resistively dividing the internal 2.5-V reference to set appropriate voltages on the inputs to the comparators.

One comparator is used in this circuit to turn on the heating element when the temperature drops below 30°C; it corresponds to a voltage of 1.52 V on the comparator's input.

MANUAL CONTROL FOR HEATER



ELECTRONIC DESIGN

Fig. 91-3

Built around an LM339 quad comparator, this circuit provides manual control of the output of a resistive heater or other load with a long time constant. The circuit's design uses minimal parts, thus it's inexpensive, and generates very low RFI.

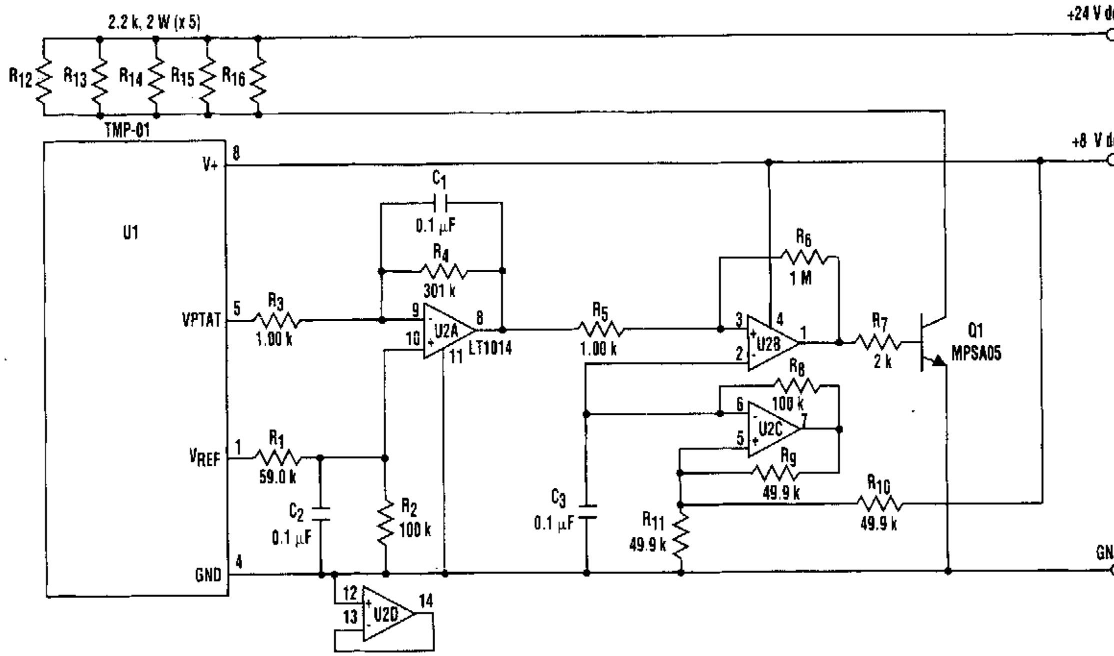
Comparators a, b, and c form a low-frequency pulse-width modulator. Sections a and b form a sawtooth oscillator (of approximately 0.25 Hz), with capacitor C1 being charged through R1 and discharged through section a's open collector output. R2 and R3 set the upper voltage limit for the sawtooth wave. The hysteresis means that C1 is discharged to nearly 0 V, creating a voltage swing that is identical to the adjustment range of R3.

Comparator c, in conjunction with potentiometer R3, converts the sawtooth wave form to a variable duty-cycle drive for the silicon-controlled rectifier.

Increasing voltage at R3's wiper means increasing the "on" time. Section d holds the SCR gate low if the line voltage is above approximately 3.5 V, preventing turn on at mid-cycle and ensuring low RFI.

The oscillator frequency is roughly determined by $1/0.7R_1C_1$. Resistance R_1 must be greater than $4R_2$ or the oscillator will lock up. Reducing R_2 will increase the lower voltage limit of the sawtooth; increasing it might cause lock-up.

PROPORTIONAL TEMPERATURE CONTROLLER



ELECTRONIC DESIGN *Fig. 91-4*

Most temperature-controller circuits use upper and lower trip points to control a heater element, with the heater power full on and full off. Usually, this results in a temperature hysteresis of several degrees. This relatively large temperature hysteresis effect might cause modulation in the output of the circuit that's being controlled.

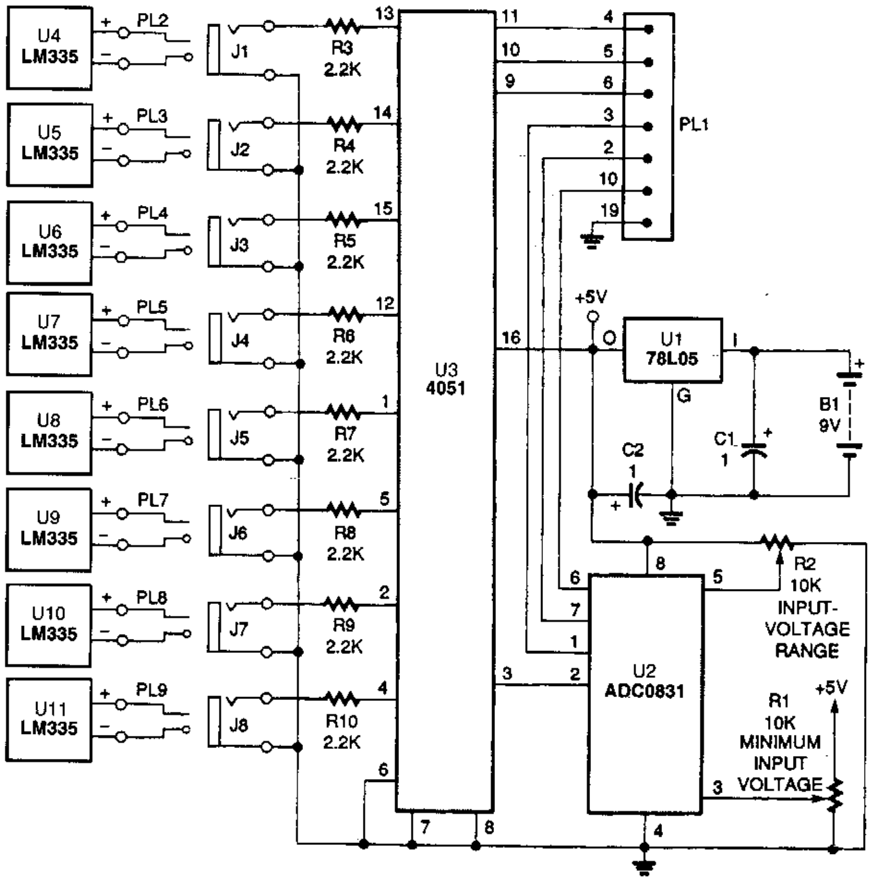
A proportional temperature controller eliminates this problem by continuously providing the power needed to maintain the "oven" at the desired temperature—within 1°C. From a cold start, maximum power is applied until the temperature is within 2°C of the set point.

The circuit's mechanical construction is important. The five heater resistors (R12 through R16), the temperature-sensor IC (U1), and the circuit being controlled are mounted with thermal epoxy to a small piece of aluminum. This provides excellent heat transfer between the components. The heater resistors must be selected to raise the temperature from ambient to the set point within an acceptable warm-up time.

U1 is Analog Devices' TMP-01 temperature-controller IC. The voltage proportional to absolute temperature (VPTAT) has a temperature coefficient of exactly 5 mV/°C. The set point is determined by the R_1/R_2 ratio. U2 is a Linear Technology LT1014 quad precision op amp. U2C is an oscillator with a 50% duty cycle that supplies a triangle wave between ⅓ and ⅔ of the supply voltage at U2-2.

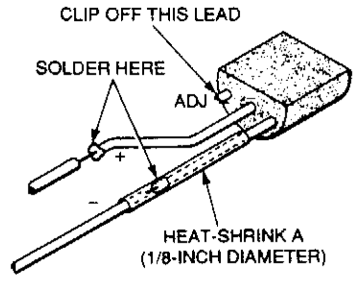
U2B compares the amplified VPTAT to the triangle wave, which drives Q1 at a duty cycle of 100% or less. Because the triangle wave's peak-to-peak amplitude is 2.7 V, and VPTAT is amplified by a factor of 300, a temperature change of approximately 2 mV moves the duty cycle from 100% to 0%.

EIGHT-INPUT A/D CONVERTER FOR TEMPERATURE MEASUREMENTS

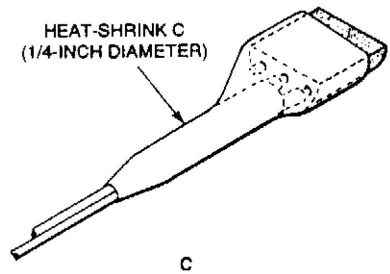


POPULAR ELECTRONICS *Fig. 91-5*

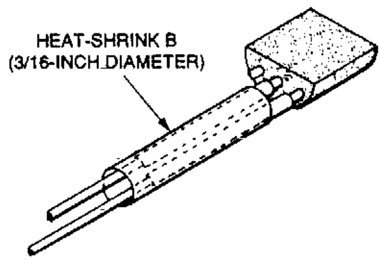
The actual processing circuitry of this A/D converter consists of only four parts: U2, U3, R1 and R2. Eight temperature probes are used with the circuit; however, they can be replaced with other types of sensors, as long as resistors R3 through R10 are removed.



A



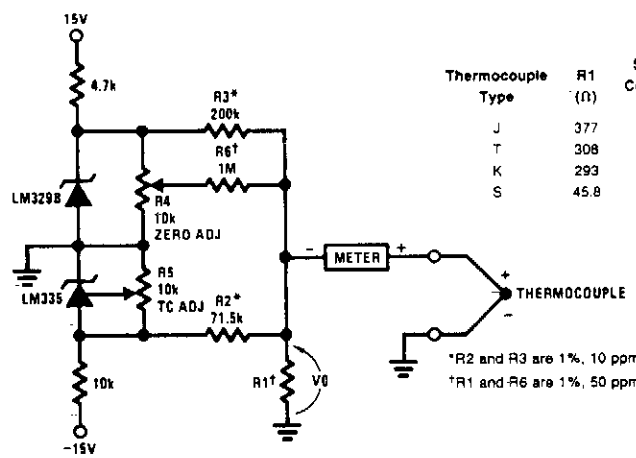
C



B

To protect the temperature probe connections, three pieces of heat-shrink tubing must be used. Heat-shrink A prevents the negative lead form coming into contact with the positve one (A). heat-shrink B protects both leads (B), and heat-shrink C covers the entire probe (C).

COLD-JUNCTION COMPENSATION FOR A GROUNDED THERMOCOUPLE



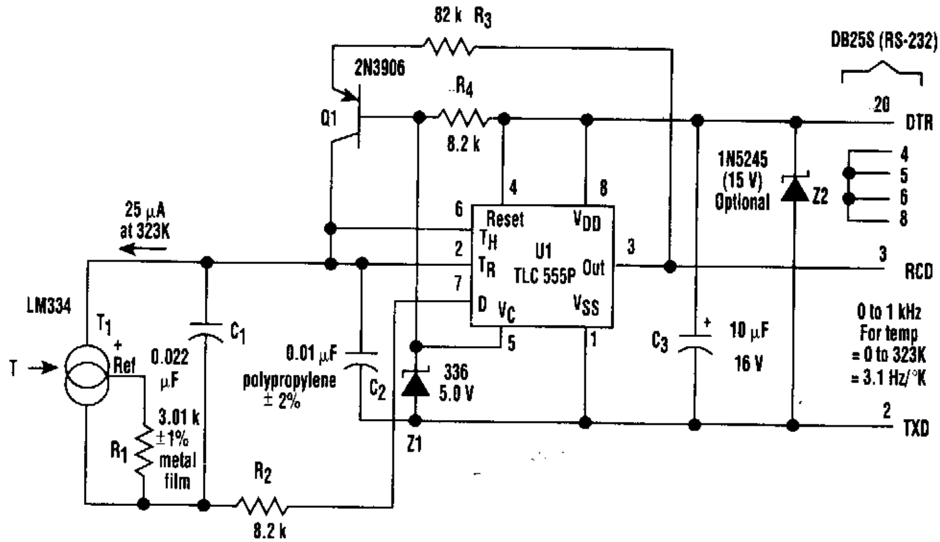
Thermocouple Type	R1 (Ω)	Seebeck Coefficient (μV/°C)
J	377	52.3
T	308	42.8
K	293	40.8
S	45.8	6.4

*R2 and R3 are 1%, 10 ppm/°C. (20 ppm/°C tracking.)
†R1 and R6 are 1%, 50 ppm/°C.

NATIONAL SEMICONDUCTOR ***Fig. 91-6***

A circuit for use with grounded thermocouples is shown. To trim, short out the LM329B and adjust R5 so that $V^o = \alpha T$, where α is the Seebeck coefficient of the thermocouple and T is the absolute temperature. Remove the short and adjust R4 so that V^o equals the thermocouple output voltage at ambient. A good grounding system is essential here, for any ground differential will appear in series with the thermocouple output.

ABSOLUTE TEMPERATURE LOG WITH RS-232



ELECTRONIC DESIGN

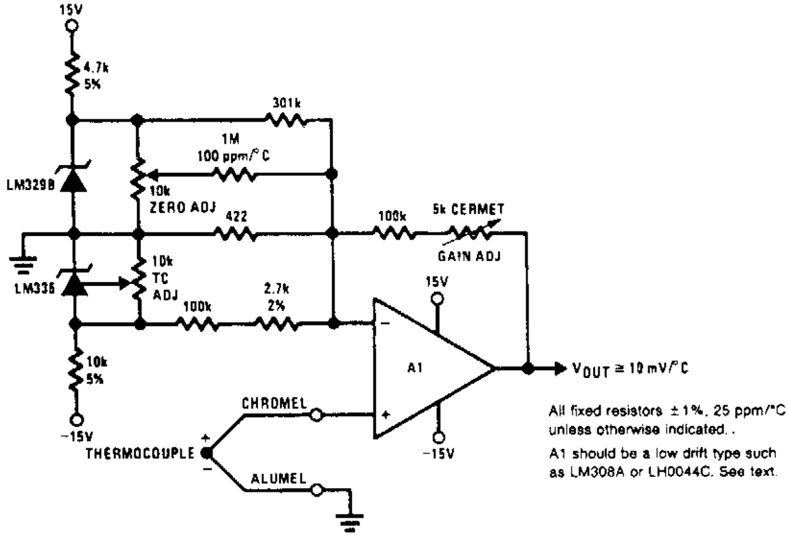
Fig. 91-7

In the setup, T1 (an LM334 temperature sensor) generates a constant current that's proportional to absolute temperature, and equal to 25 μA at 323 Kelvin (50°C). R1 sets this constant of proportionality. The current discharges the parallel combination of C1 and C2 connected to the trigger and the threshold pins of U1, which is a CMOS implementation of the venerable 555 analog timer. The negative-going ramp is compared by U1 to an internal 2.5-V trigger level controlled by Z1. When the ramp gets there, U1 triggers. The output pin (3) will go high, presenting a "start" bit to the connected communications port, which causes Q1 to source 100 μA to the timing node. At the same time, the discharge pin (7) will open, allowing R2 and the bottom end of C1 and T1 to be supplied by C1, and isolates it from C2.

Consequently, the current supplied by transistor Q1 will go solely to C2 so that when the resulting positive-going ramp reaches 5 V, exactly 25 nC (2.5 V × C2) will have been deposited in the timing node by the recharge cycle because its threshold level (6) will have been reached. This causes both the output pin to return to the negative rail, restoring the "marking" condition of the R2-232 interface, and Q1 to stop recharging C2. U1's discharge pin (7) now connects R2 to the negative rail, causing the charge that was deposited by T1 on C1 during the recharge interval to rapidly redistribute between C1 and C2. This arrangement creates a very linear (0.01%) relationship between T1 current and pulse output frequency.

While this is happening, the PC's communications port hardware assembles a valid (although meaningless) character because the positive pulse output by U1 looks like the start bit of a character. A simple program running in the PC can then count the frequency of these characters and convert the resulting rate into a direct readout of temperature. Because of the 3.1-Hz/degree slope of frequency versus temperature, a 30-second average suffices for 0.01° resolution.

CENTIGRADE THERMOMETER WITH COLD-JUNCTION COMPENSATION

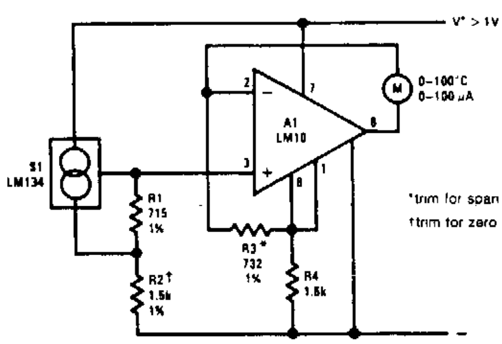


NATIONAL SEMICONDUCTOR

Fig. 91-8

This electronic thermometer has a 10-mV/°C output from 0°C to 1300°C. The trimming procedure is as follows: first short out the LM329B, the LM335 and the thermocouple. Measure the output voltage (equal to the input offset voltage times the voltage gain). Then apply a 50-mV input voltage and adjust the GAIN ADJUST pot until the output voltage is 12.25 V above the previously measured value. Next, short out the thermocouple again and remove the short across the LM335. Adjust the TC ADJUST pot so that the output equals 10 mV/°K times the absolute temperature. Finally, remove the short across the LM329B and adjust the ZERO ADJUST pot so that the output voltage equals 10 mV/°C times the ambient temperature in °C.

1.5-V ELECTRONIC THERMOMETER

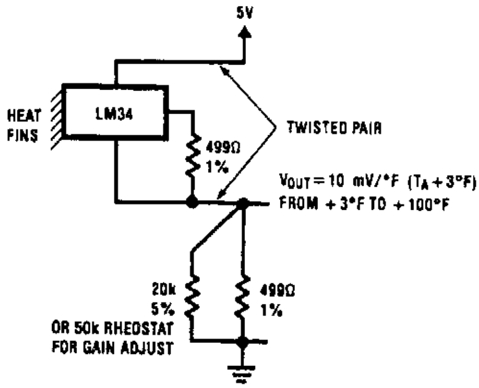


NATIONAL SEMICONDUCTOR

Fig. 91-9

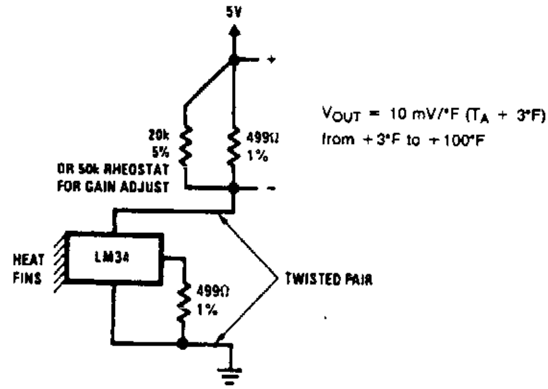
An electronic thermometer design, useful in the range of –55°C to 150°C, is shown. The sensor, S1, develops a current that is proportional to absolute temperature. This is given the required offset and range expansion by the reference and op amp, resulting in a direct readout in either °C or °F.

TWO-WIRE TEMPERATURE SENSOR OUTPUT REFERENCED TO GROUND



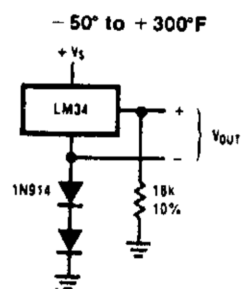
NATIONAL SEMICONDUCTOR *Fig. 91-10*

TWO-WIRE REMOTE TEMPERATURE SENSOR WITH SENSOR GROUNDED



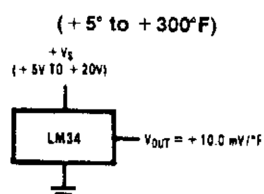
NATIONAL SEMICONDUCTOR *Fig. 91-11*

SINGLE-SUPPLY TEMPERATURE SENSOR (−50 TO +300°F)



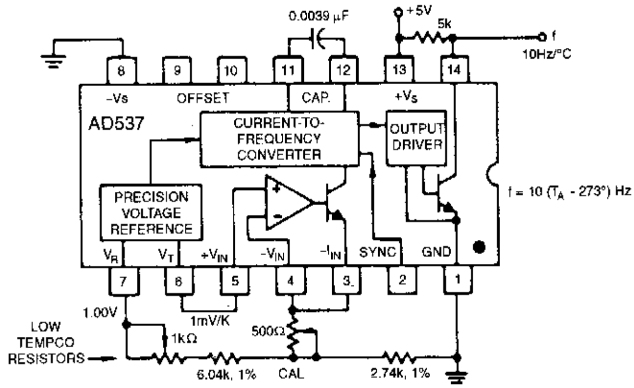
NATIONAL SEMICONDUCTOR *Fig. 91-12*

BASIC FAHRENHEIT TEMPERATURE SENSOR



NATIONAL SEMICONDUCTOR *Fig. 91-13*

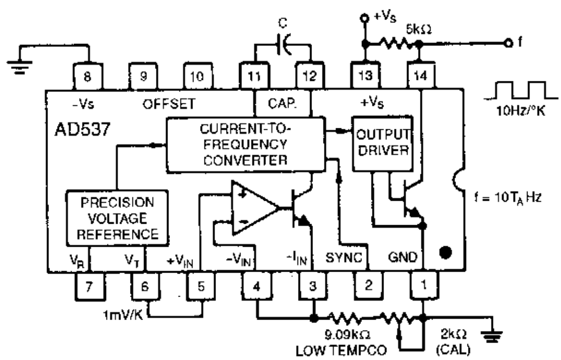
TEMPERATURE-TO-FREQUENCY CONVERTER (CELSIUS)



ANALOG DEVICES *Fig. 91-14*

The 1.00-V reference output can be combined with the 1-mV/°K output to realize various temperature scales. For the Celsius scale, the lower end of the timing resistor must be offset by 273 mV. This is easily accomplished, and it results in an output from 0 to 1 kHz for temperatures from 0°C to +100°C. Other offsets and scale factors are equally easy to implement.

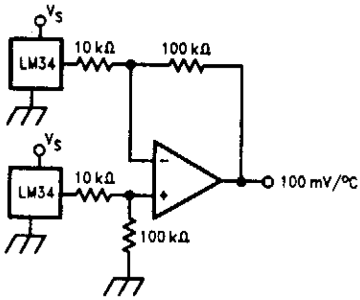
TEMPERATURE-TO-FREQUENCY CONVERTER (KELVIN)



ANALOG DEVICES *Fig. 91-15*

This simple connection results in a direct conversion of temperature to frequency. The 1-mV/°K temperature output serves as the input to the buffer amplifier, and the oscillator drive current is scaled to be 298 μA at 298°K (+25°C). Use of a 1000-pF capacitor results in a corresponding frequency of 2.98 kHz. A single-point trim for calibration is normally sufficient to give errors less than ±2°C from –55°C to +125°C. An NPO capacitor is preferred to minimize nonlinearity that results from capacitance drift.

DIFFERENTIAL THERMOMETER

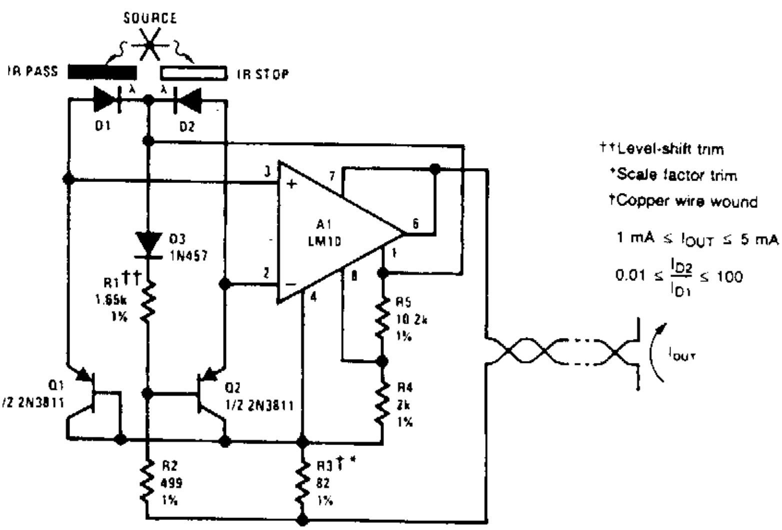


The differential thermometer shown in the figure produces an output voltage which is proportional to the temperature difference between two sensors. This is accomplished by using a difference amplifier to subtract the sensor outputs from one another and then multiplying the difference by a factor of 10 to provide a single-ended output of 100 mV per degree of differential temperature.

NATIONAL SEMICONDUCTOR

Fig. 91-16

OPTOELECTRONIC PYROMETER

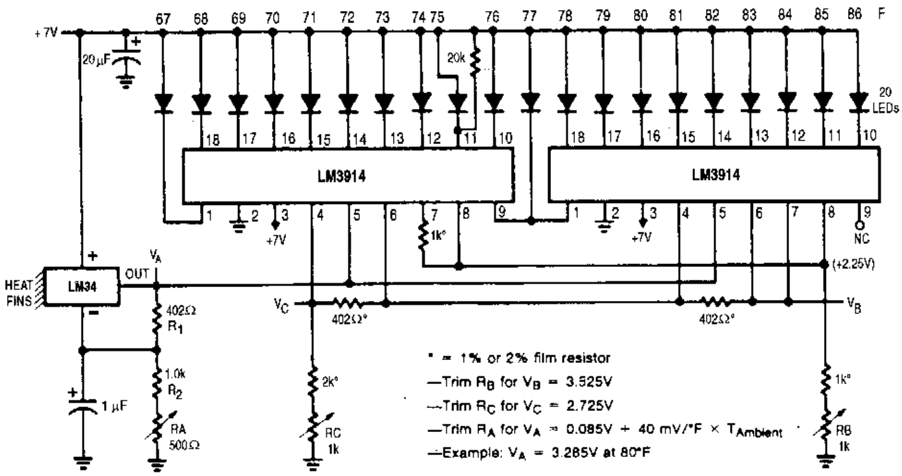


NATIONAL SEMICONDUCTOR

Fig. 91-17

This setup optically measures the temperature of an incandescent body. It makes use of the shift in the emission spectrum of a black body toward shorter wavelengths, as temperature is increased. Optical filters are used to split the emission spectrum, with one photodiode being illuminated by short wavelengths (visible light) and the other by long (infrared). The photocurrents are converted to logarithms by Q1 and Q2. These are subtracted to generate an output that varies as the log of the ratio of the illumination intensifies. Thus, the circuit is sensitive to changes in spectral distribution, but not intensity.

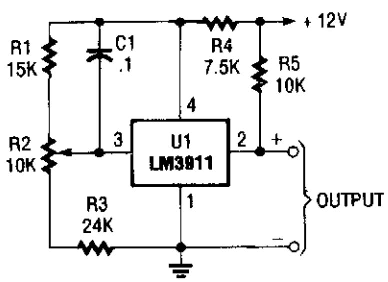
BAR-GRAPH ROOM TEMPERATURE DISPLAY



NATIONAL SEMICONDUCTOR *Fig. 91-18*

This display shows temperature as a bar graph. The range is 67°F to 86°F.

LM3911 TEMPERATURE CONTROLLER



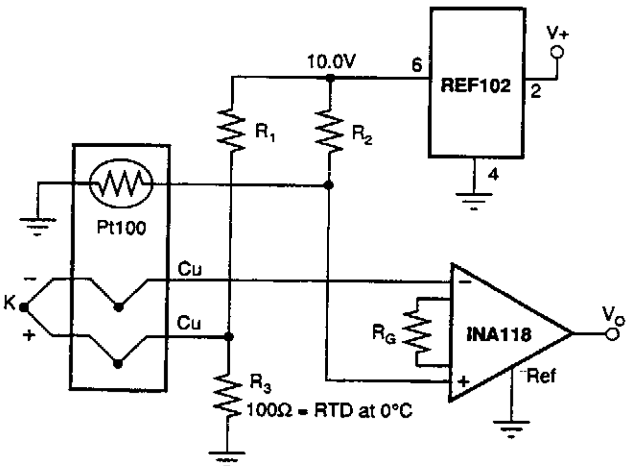
POPULAR ELECTRONICS *Fig. 91-19*

The LM3911 requires only a few resistors and a capacitor to turn into a full-function temperature controller. The LM3911 requires very few external components to implement a full-function temperature controller. The figure illustrates the simplicity of the freeze fighter circuit. The resistor network consisting of R1, R2, and R3 is used to provide the set point voltage for the feedback input of the LM3911.

Resistor R4 limits the current through the internal voltage reference of the LM3911 and can be selected from a wide range of values. 7.5 kΩ is specified in most of the application notes.

Resistor R5 pulls the output of the IC high when the temperature is below the set point. The internal resistor is in series with a diode that allows a switching voltage up to 35 V to be used with the device.

THERMOCOUPLE AMPLIFIER WITH COLD-JUNCTION COMPENSATION



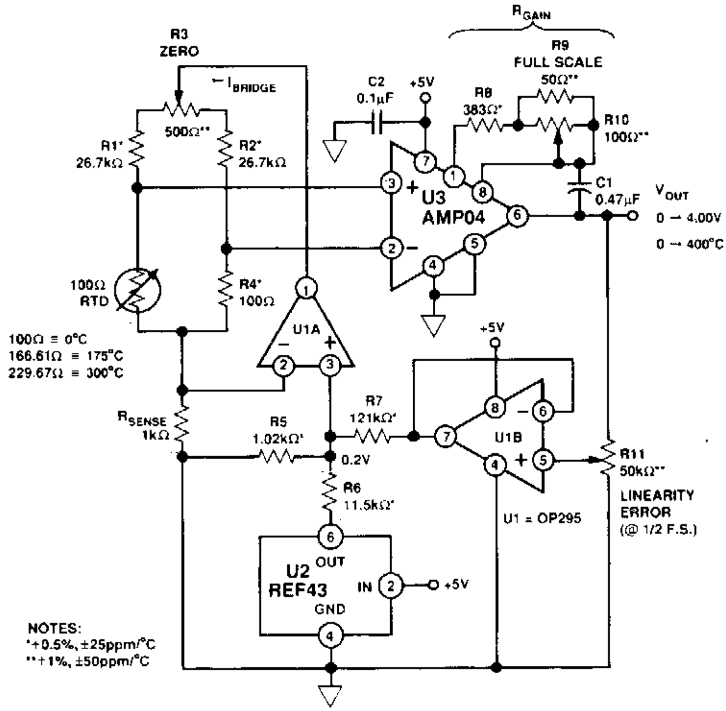
ISA TYPE	MATERIAL	SEEBECK COEFFICIENT (μV/°C)	R_1, R_2
E	+ Chromel − Constantan	58.5	66.5kΩ
J	+ Iron − Constantan	50.2	76.8kΩ
K	+ Chromel − Alumel	39.4	97.6kΩ
T	+ Copper − Constantan	38.0	102kΩ

DESIRED GAIN	R_G (Ω)	NEAREST 1% R_G (Ω)
1	NC	NC
2	50.00k	49.9k
5	12.50k	12.4k
10	5.556k	5.62k
20	2.632k	2.61k
50	1.02k	1.02k
100	505.1	511
200	251.3	249
500	100.2	100
1000	50.05	49.9
2000	25.01	24.9
5000	10.00	10
10000	5.001	4.99

BURR-BROWN

Fig. 91-20

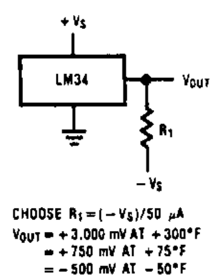
PRECISION RTD AMPLIFIER CIRCUIT FOR +5 V



ANALOG DEVICES *Fig. 91-21*

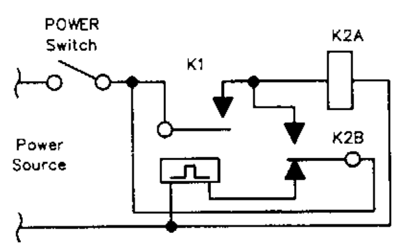
This circuit uses a platinum resistance temperature device to sense temperature. It has a range of 0 to 300°C. The RTD bridge is driven with a regulated 200-μA current to minimize self heating of the RTD. A 5-V supply is used.

FULL-RANGE FAHRENHEIT TEMPERATURE SENSOR



NATIONAL SEMICONDUCTOR *Fig. 91-22*

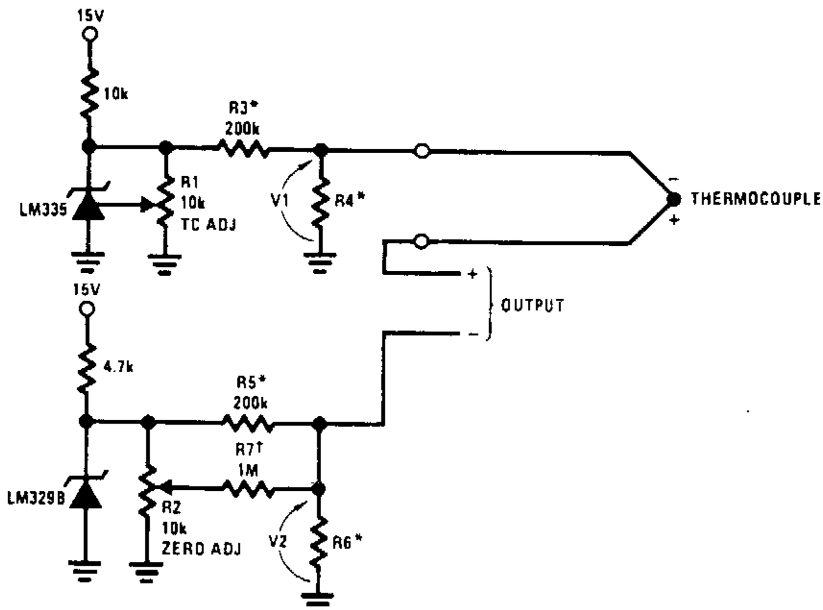
IMPROVED THERMOSTATIC RELAY CIRCUIT



(Additional Circuit Switched by K2 not Shown)

QST *Fig. 91-23*

THERMOCOUPLE COLD-JUNCTION COMPENSATION



Thermocouple Type	Seebeck Coefficient (μV/°C)	R4 (Ω)	R6 (Ω)
J	52.3	1050	385
T	42.8	856	315
K	40.8	816	300
S	6.4	128	46.3

*R3 thru R6 are 1%, 5 ppm/°C. (10 ppm/°C tracking.)

†R7 is 1%, 25 ppm/°C.

choose $R4 = \frac{\alpha}{10\ mV/^{\circ}C} \cdot R3$

choose $R6 = \frac{-T_o \cdot \alpha}{V_Z} \cdot (0.9R5)$

choose $R7 = 5 \cdot R5$

where T_o is absolute zero (−273.16°C)

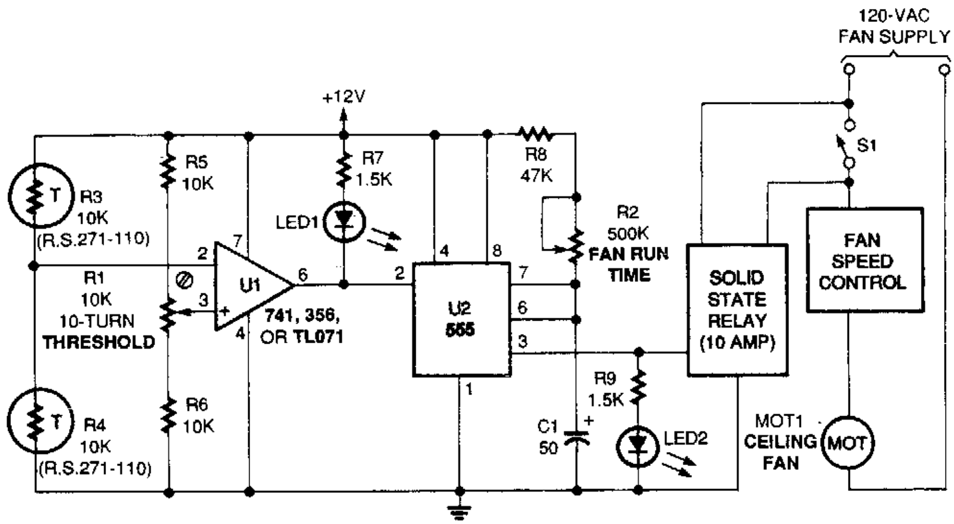
V_Z is the reference voltage (6.95V for LM329B)

NATIONAL SEMICONDUCTOR

Fig. 91-24

A single-supply circuit is shown. R3 and R4 divide down the 10-mV/°K output of the LM329B and its associated voltage divider provide a voltage to buck out the 0°C output of the LM335. To calibrate, adjust R_1 so that $V_1 = \alpha\ T$, where α is the Seebeck coefficient and T is the ambient temperature in degrees Kelvin. Then, adjust R_2 so that $V_1 - V_2$ is equal to the thermocouple output voltage at the known ambient temperature.

TEMPERATURE DIFFERENTIAL DETECTOR

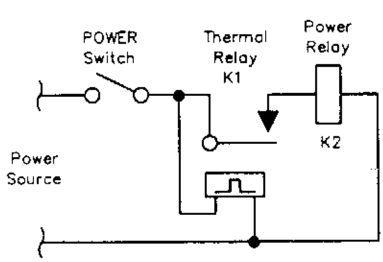


POPULAR ELECTRONICS *Fig. 91-25*

This circuit measures temperature differences, not temperature. Once the difference passes a certain threshold, the timer is triggered, activating the solid-state relay.

Op amp U1 is placed in a comparator configuration with two thermistors—one located at the ceiling, one at the floor. The IC senses the temperature difference between the ceiling and floor, but is unaffected by the overall room temperature differential increases. The upper thermistor will decrease in resistance, eventually causing the voltage at pin 2 of U1 to exceed that pin 2 of U1 to go low and trip the timer. This adds hysteresis to the op amp's output, preventing the motor from chattering. When tripped, the timer activates the relay, which activates the fan.

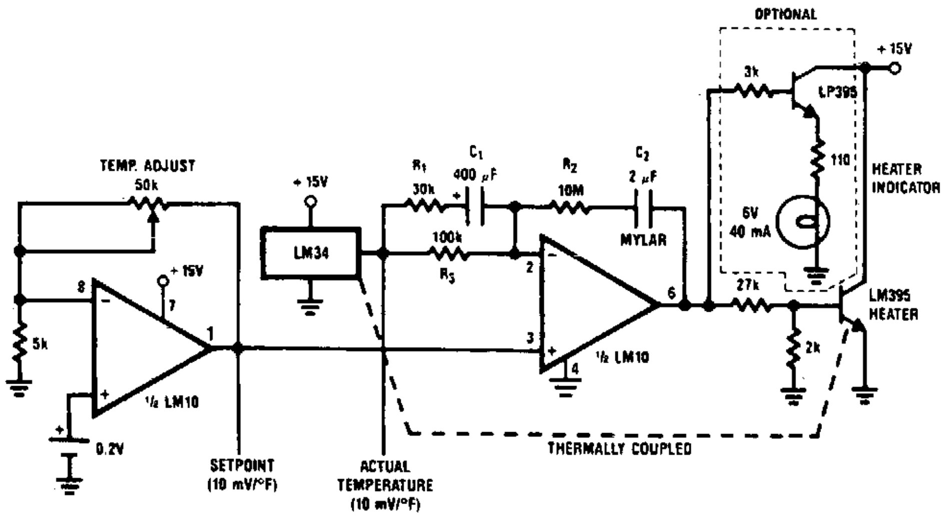
THERMOSTATIC RELAY APPLICATION



K1, the thermostatic relay, energizes power relay K2, which handles the circuit's power switching. The drawing doesn't show K2's power-switching contacts.

QST *Fig. 91-26*

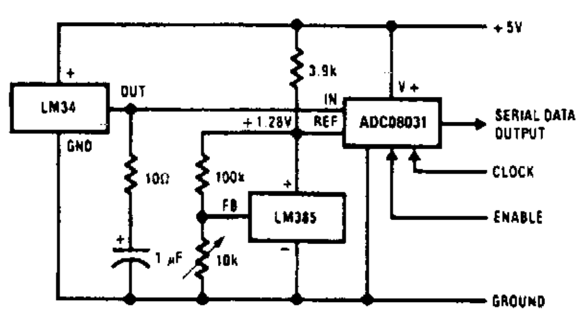
TEMPERATURE CONTROLLER



NATIONAL SEMICONDUCTOR *Fig. 91-27*

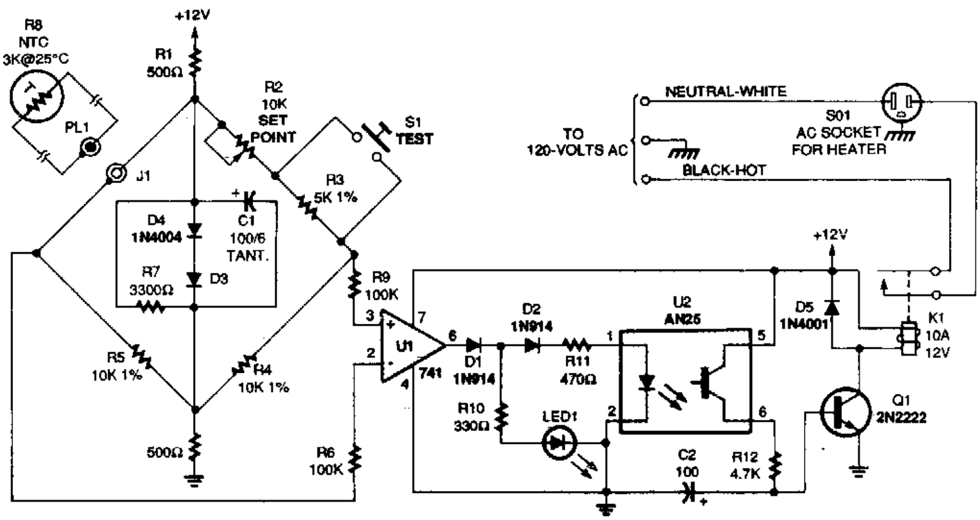
A proportional temperature controller can be made with an LM34 and a few additional parts. The complete circuit is shown. Here, an LM10 serves as both a temperature-setting device and as a driver for the heating unit (an LM395 power transistor). The optional lamp, driven by an LP395 transistor, is for indicating whether or not power is being applied to the heater.

TEMPERATURE-TO-DIGITAL-OUTPUT CONVERTER



NATIONAL SEMICONDUCTOR *Fig. 91-28*

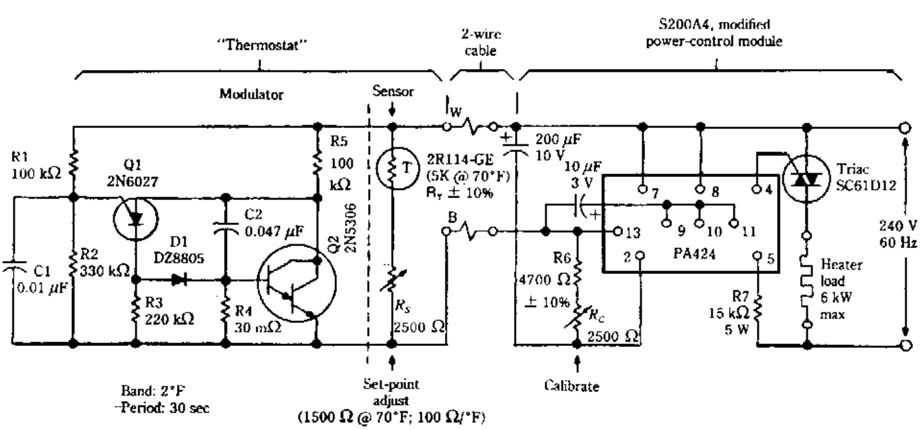
FREEZE-UP SENSOR



POPULAR ELECTRONICS ***Fig. 91-29***

Using a bridge circuit to provide an accurate activation temperature, this circuit will turn on a heating unit or other device when the temperature drops below the trip point set by R2. Use a 10-kΩ resistor in place of the thermistor to calibrate it for 32°F activation.

ZERO-VOLTAGE SWITCHING TEMPERATURE REGULATOR



McGRAW-HILL ***Fig. 91-30***

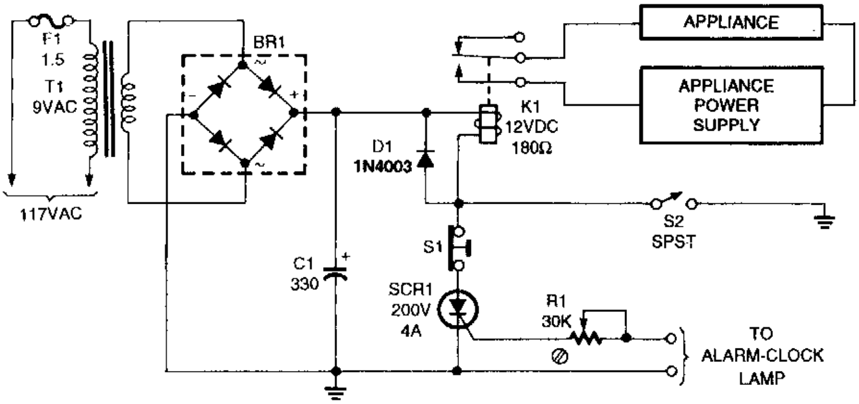
In this arrangement, an integral number of cycles of ac is fed to the heater. No RFI or EMI is generated with this method. The thermostat uses a thermistor as a sensor. The PA424 (GE) device generates trigger pulses for the triac only at zero crossings of the ac line cycle.

92

Timer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

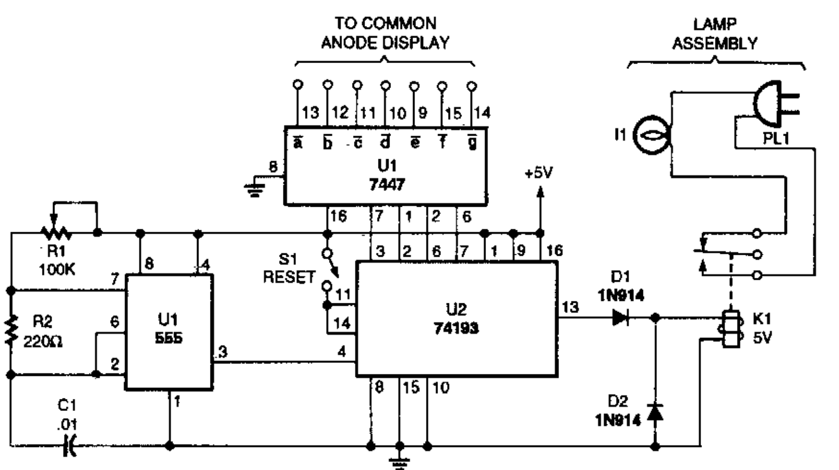
ALARM CLOCK TIMER



POPULAR ELECTRONICS *Fig. 92-1*

Turn your alarm clock into a specialized timer with this simple circuit. The clock used with the circuit should be the kind that turns on a little lamp when the alarm is activated.

LAMP TIMER

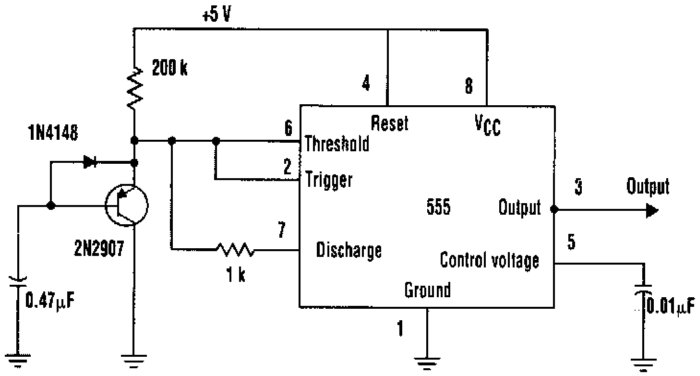


POPULAR ELECTRONICS *Fig. 92-2*

A timed switch uses a 555 oscillator/timer wired to operate in the astable mode. The timer supplies a positive pulse to the clock input of a 74193 4-bit binary up/down count every five minutes. Because the 74193 is set to operate in the count-down mode, the output of the 555 is connected to the count-down input of the 74193.

As the binary counter is reset, it starts counting at nine and counts down to zero with each clock pulse. When the counter hits zero, the output from the 74193 goes low, turning off the relay and the light. The light can be turned back on by pressing the reset button again.

LONG-PERIOD TIMER



ELECTRONIC DESIGN *Fig. 92-3*

Adding a transistor to the 555 timer can create long timer periods, which is a key factor when the timer is operating at low speed. The transistor basically acts as a current divider or capacitance multiplier. The problem with low speed, however, is that the timing resistors and capacitors must be large and the charging current must be small, particularly when the desired timing period is in the range of seconds.

Typically, electrolytic capacitors are used in these situations, but their leakage current tends to aggravate or even prohibit operation at very low charging currents.

This problem can be solved by adding a transistor. In effect, the transistor is used as a current divider or a capacitance multiplier. The normal charging current (emitter current) is divided by the transistor's current gain so that the capacitor charging current (base current) is reduced considerably. For example, 10 μA of emitter current will require approximately 0.1 μA of base current, based on a current gain of 100.

In this circuit, the capacitor will be charged with such a low charging current that timing periods will typically be 100 times longer than usual. This means that substantial time periods can be achieved with film or ceramic capacitors that have much better leakage characteristics and are physically smaller.

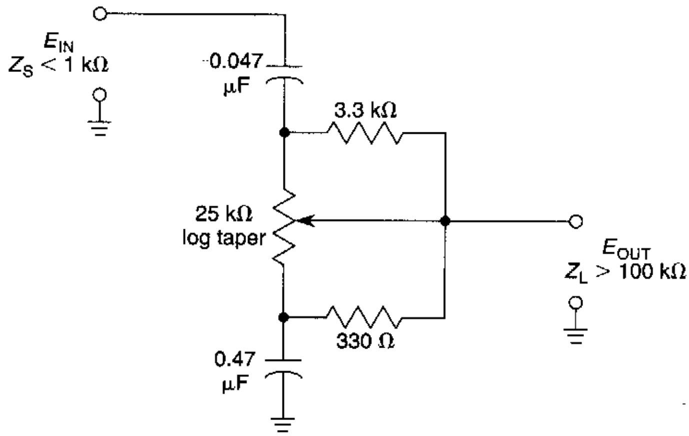
The circuit's output period was approximately 6 seconds, compared to 80 ms without the transistor. The transistor multiplied the normal time period by a factor of approximately 75.

93

Tone Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

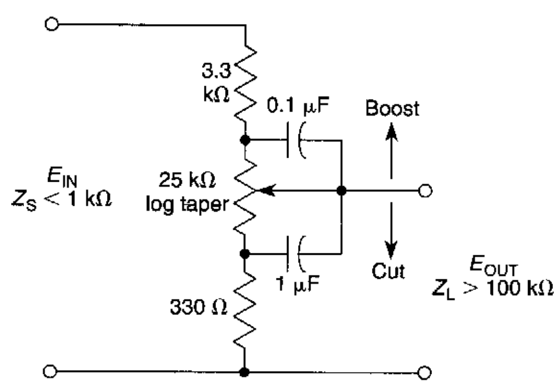
TREBLE-CONTROL CIRCUIT



WILLIAM SHEETS ***Fig. 93-1***

This tone control has an insertion loss of 20 dB at flat setting and is effective above 1 kHz. It has little effect below about 1 kHz.

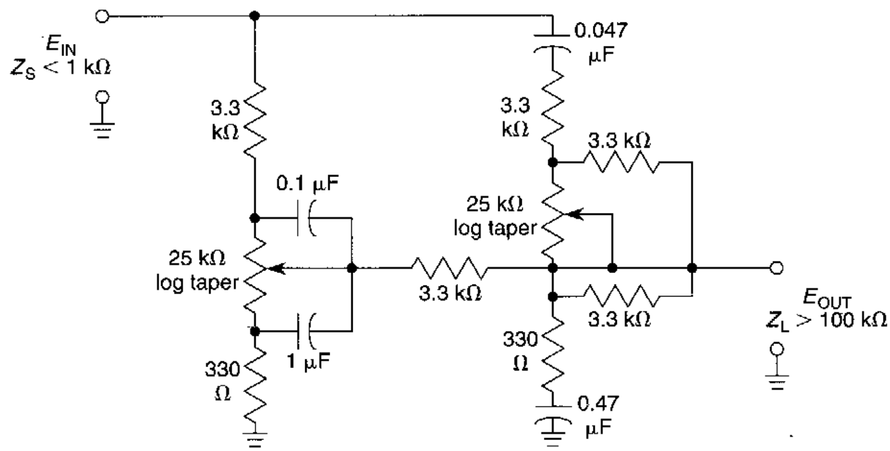
BASS TONE-CONTROL CIRCUIT



WILLIAM SHEETS ***Fig. 93-2***

This tone control has an insertion loss of 20 dB at flat setting and is effective below 350 Hz. The control has little effect above this frequency.

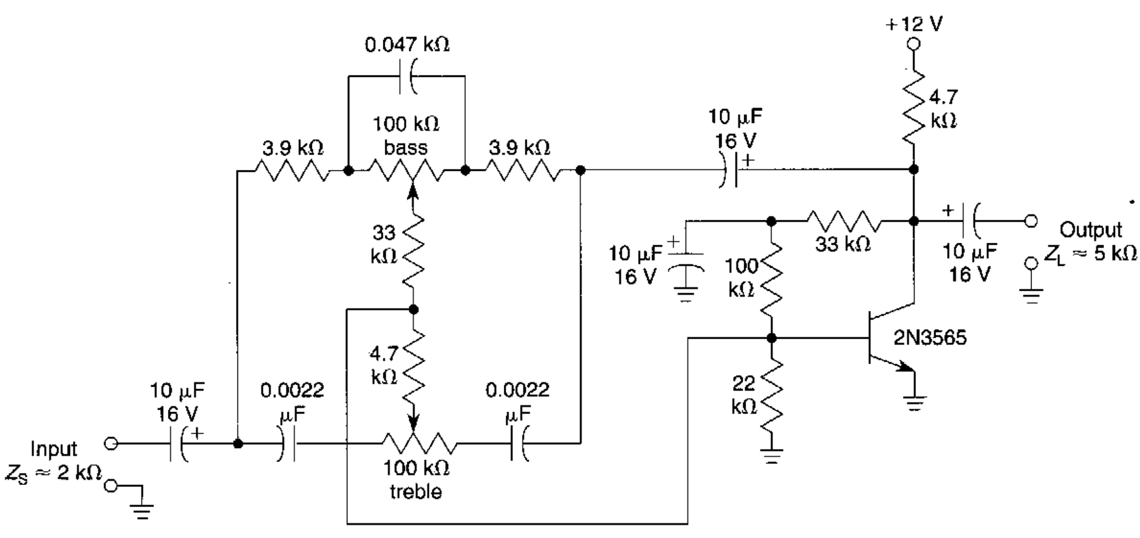
COMBINED BASS AND TREBLE CONTROL



WILLIAM SHEETS ***Fig. 93-3***

This positive tone control system uses two pots to control bass and treble.

ACTIVE BASS- AND TREBLE-TONE CONTROL



WILLIAM SHEETS ***Fig. 93-4***

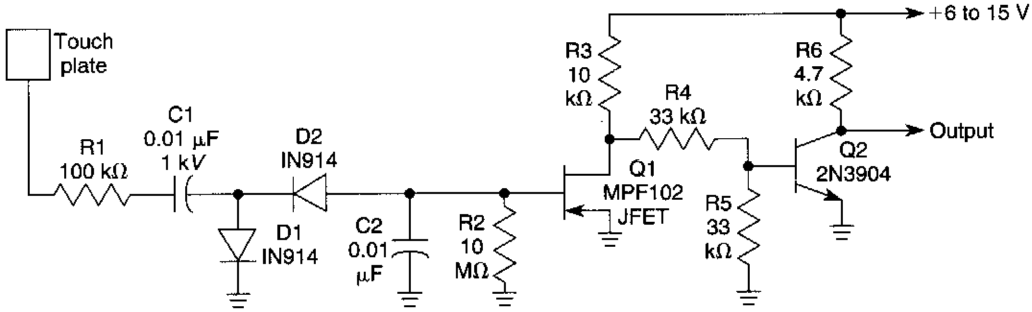
A single transistor used as a feedback amplifier is connected with ac feedback through the tone controls, which determine the frequency response of the stage.

94

Touch/Proximity Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE TOUCH SWITCH

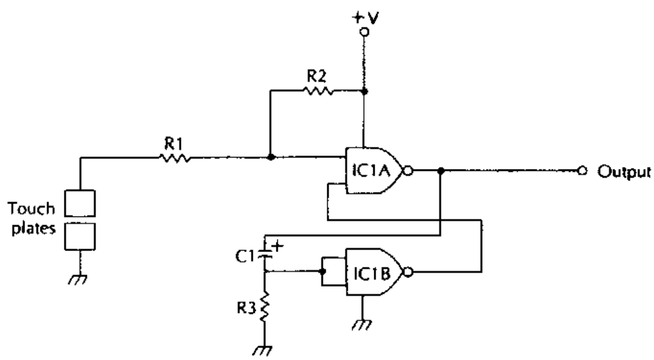


WILLIAM SHEETS

Fig. 94-1

Q2 is held cut off since Q1 normally is conducting. When the touch plate is contacted by a large object (human body, etc.), stray 60-Hz pickup is rectified by D1 and D2, and produces a negative voltage across R2-C2 and the gate of Q1. Q1 cuts off, causes Q2 to conduct, and the output goes low.

SIMPLE TIMED TOUCH SWITCH



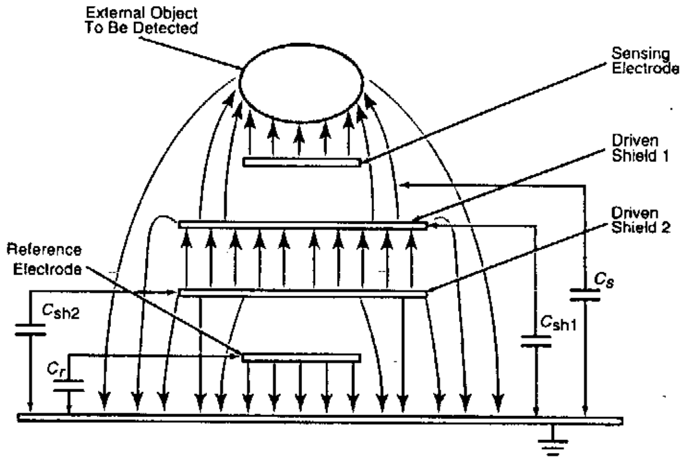
McGRAW-HILL

Fig. 94-2

This circuit produces an output for a time approximately equal to time constant R_3C_1.

IC1, IC2	CD4011 quad NAND gate
IC3	CD4066 quad bilateral switch
C1	47-µF, 25-V electrolytic capacitor
C2	100-µF, 25-V electrolytic capacitor
C3	220-µF, 25-V electrolytic capacitor
C4	470-µF, 25-V electrolytic capacitor
C5, C6, C7	0.1-µF capacitor
R1, R3, R4, R6 R7, R9, R10, R12	100-kΩ, ¼-W 5% resistor
R2, R5, R8, R11	10-MΩ, ¼-W 5% resistor

CAPACITIVE SENSOR SYSTEM



NASA TECH BRIEFS ***Fig. 94-3***

This figure illustrates the electric-field configuration of a capacitive proximity sensor of the "capaciflector" type. It includes a sensing electrode driven by an alternating voltage, which gives rise to an electric field in the vicinity of the electrode; an object that enters the electric field can be detected by its effect on the capacitance between the sensing electrode and electrical ground.

Also, it includes a shielding electrode (in this case, driven shield 1), which is excited via a voltage follower at the same voltage as that applied to the sensing electrode to concentrate more of the electric outward from the sensing electrode, increasing the sensitivity and range of the sensor. Because the shielding electrode is driven via a voltage follower, it does not present a significant electrical load to the source of the alternating voltage.

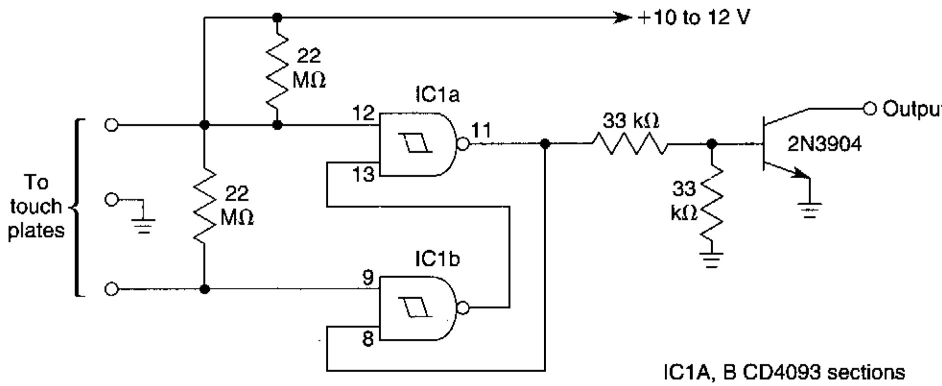
In this case, the layered electrode structure also includes a reference electrode adjacent to ground, plus a second shielding electrode (driven shield 2), which is excited via a voltage follower at the same voltage as that applied to the reference electrode. Driven shield 2 isolates the reference electrode from the electric field generated by driven shield 1 and the sensing electrode so that a nearby object exerts no capacitive effect on the reference electrode.

The excitation is supplied by a crystal-controlled oscillator and applied to the sensing and reference electrodes via a bridge circuit. Fixed capacitors C1 and C2 (or, alternatively, fixed resistors R1 and R2) are chosen to balance the bridge; that is, to make the magnitude of the voltage at sensing-electrode node S equal the magnitude of the voltage at reference-electrode node R.

The voltages at S and R are peak-detected and fed to a differential amplifier, which puts out voltage V_u proportional to the difference between them. When no object intrudes into the electric field of the sensing electrode, the bridge remains in balance, and $V_u - 0$. When an object intrudes, it changes C_s, unbalances the bridge, and causes V_u to differ from zero. The closer the object comes to the sensing electrode, the larger (V_u) becomes.

An additional output voltage KV_r is available, where K is the amplification and V_r is the voltage on the reference electrode.

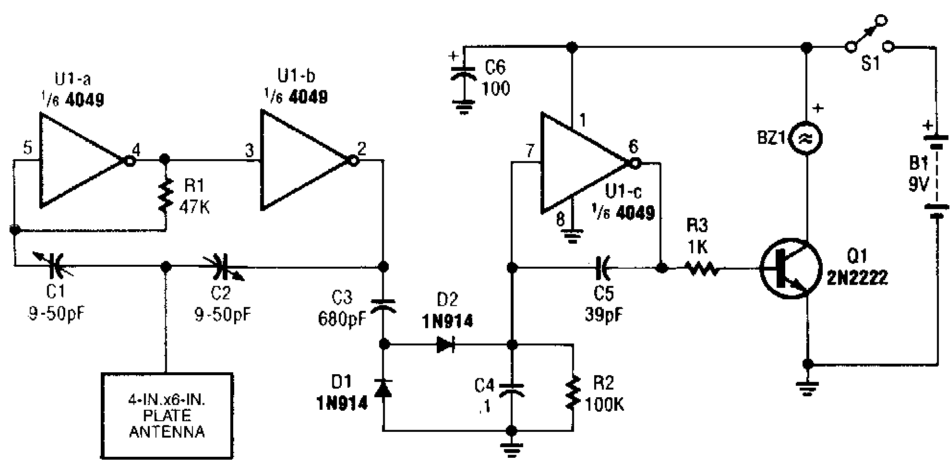
TOUCH SWITCH



WILLIAM SHEETS ***Fig. 94-4***

Two NAND Schmitt triggers are used as a flip-flop to produce a bridged touch switch.

PROXIMITY ALARM



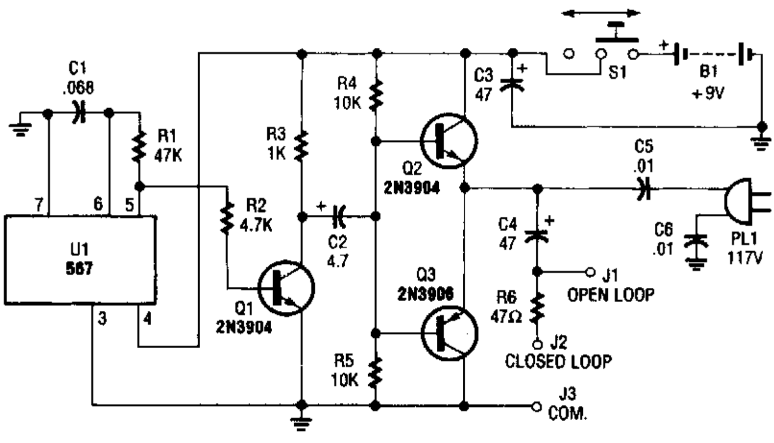
POPULAR ELECTRONICS ***Fig. 94-5***

95

Tracer Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

WIRE TRACER

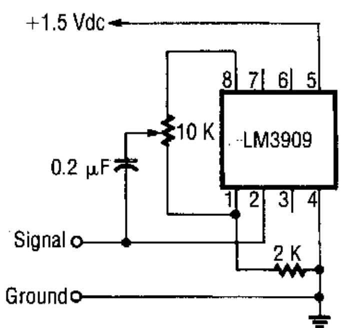


At the heart of the McTrak is a 567 tone decoder, configured as a simple squarewave oscillator, operating at about 250 Hz.

ELECTRONICS HOBBYIST HANDBOOK ***Fig. 95-1***

This tracer works by placing a square-wave signal on the line to be traced. The square wave is rich in harmonics. A small transistor radio placed close to a wire carrying this signal will buzz. The radio, therefore, is used as a probe to trace out the wire.

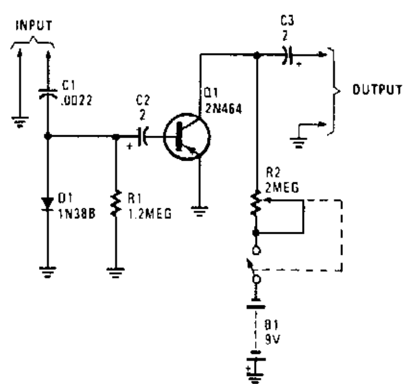
CABLE TRACER



ELECTRONICS NOW ***Fig. 95-2***

This circuit generates a 1-kHz square wave for cable tracing. Because this circuit is simple and generates from 1.5 V, several can be used at the same time to generate multiple tones for tracing multiconductor cables.

SIGNAL TRACER



POPULAR ELECTRONICS ***Fig. 95-3***

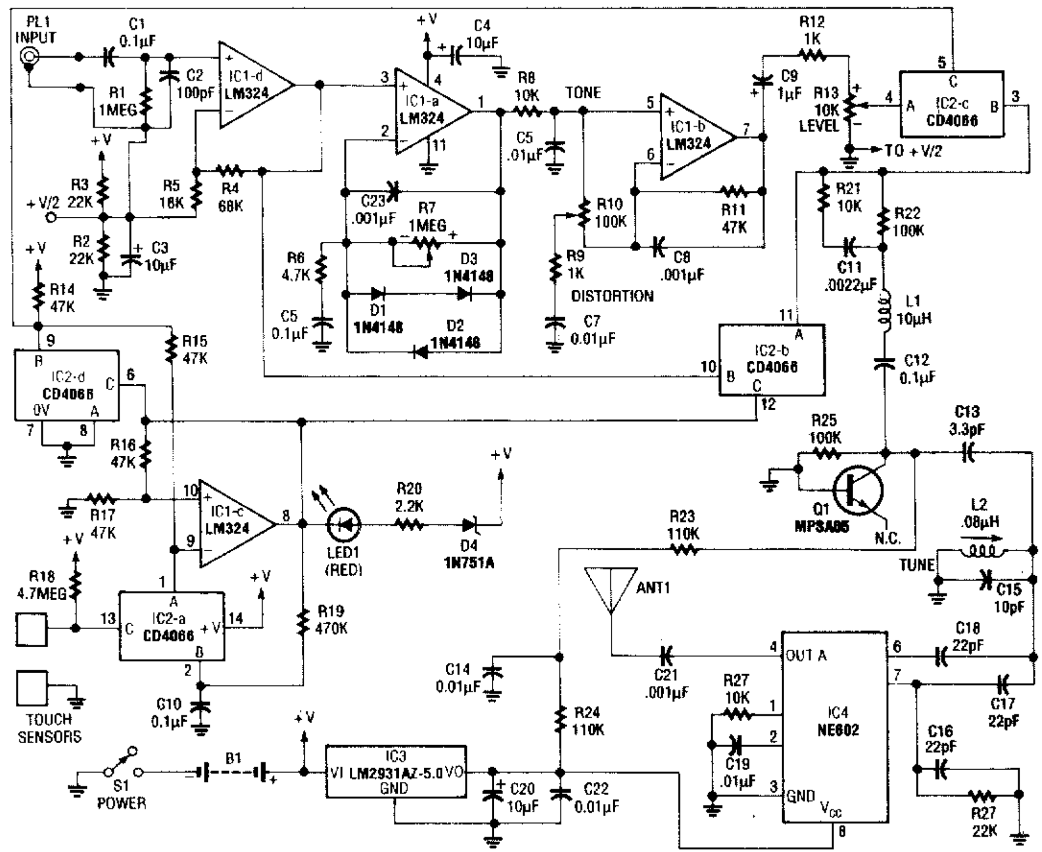
This circuit uses a simple detector-audio amplifier. The output can be connected to headphones or another audio amplifier.

96

Transmitter and Transceiver Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

WIRELESS GUITAR TRANSMITTER

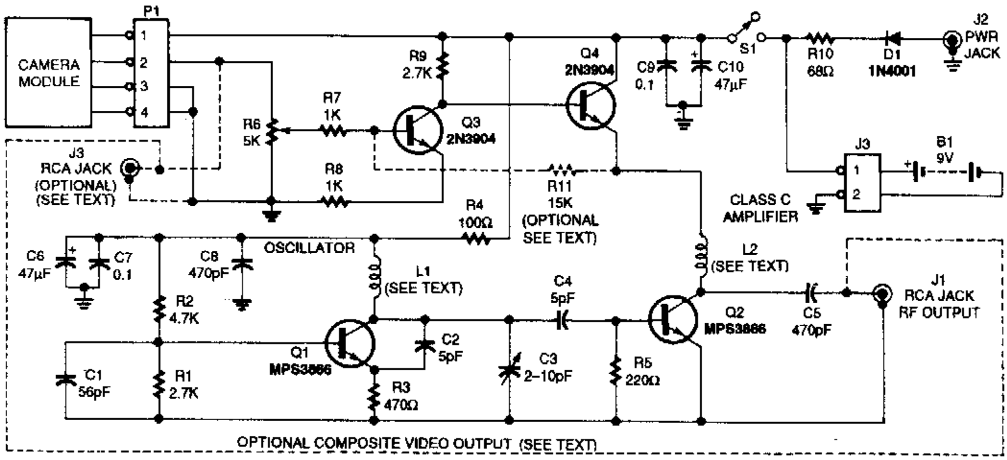


ELECTRONICS NOW

Fig. 96-1

This transmitter has a built-in distortion effects unit and a touch switch to switch effects off and on. The circuit operates from a 9-V battery. IC1-a and IC1-b are used in the effects circuitry. IC1-d is an input preamp and IC2 is a quad analog switch to handle audio switching. Q1 acts as a varactor, diode modulator while IC-4 is an 88- to 108-MHz FM oscillator.

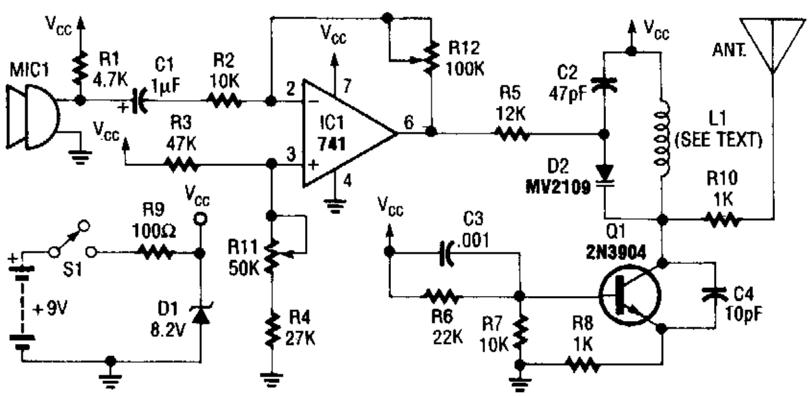
MICRO TV TRANSMITTER



ELECTRONICS NOW *Fig. 96-2*

For very low power, noncritical applications, this small TV modulator can be useful as a short-range (50 feet) transmitter for video signals. A small camera module can be used as a source. R11 is used to vary dc offset of the modulator.

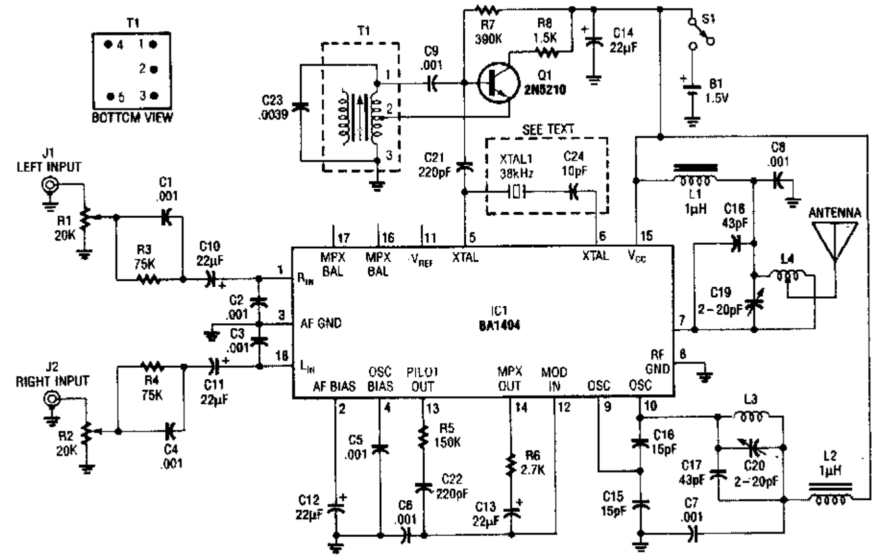
WIRELESS MICROPHONE



ELECTRONICS EXPERIMENTERS HANDBOOK *Fig. 96-3*

An op-amp IC (741) amplifies the audio signal from MIC1, and R12 controls its gain. Audio is fed to the oscillator circuit Q1 and related components. D2 is a varactor diode. Audio fed to D2 causes FM of the oscillator signal. L1 is 2½ turns of #18 wire on a 3⁄16" diameter form. The antenna is a 12" whip.

FM STEREO TRANSMITTER

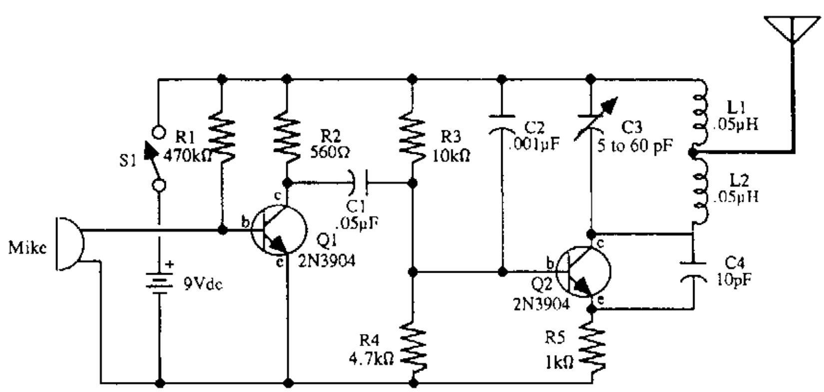


THE HEART OF THE FM TRANSMITTER is a BA1404 FM stereo transmitter IC. The left input-signal level is adjusted via R1, pre-emphasis is provided by C1 and R3, and audio is coupled by C10 into the left-channel input. The right-channel input circuitry is identical.

ELECTRONICS NOW *Fig. 96-4*

An FM stereo transmitter can be built around the BA1404 IC. This IC has all the functions necessary to generate an FM MPX signal. A separator oscillator circuit uses a 2N5210 transistor instead of the difficult-to-find 38-kHz crystal that is normally used. T1 is a 455-kHz IF transformer with 0.0039-µF capacitance added across it to enable tuning to 38 kHz. With this circuit, oscillator stability should be adequate.

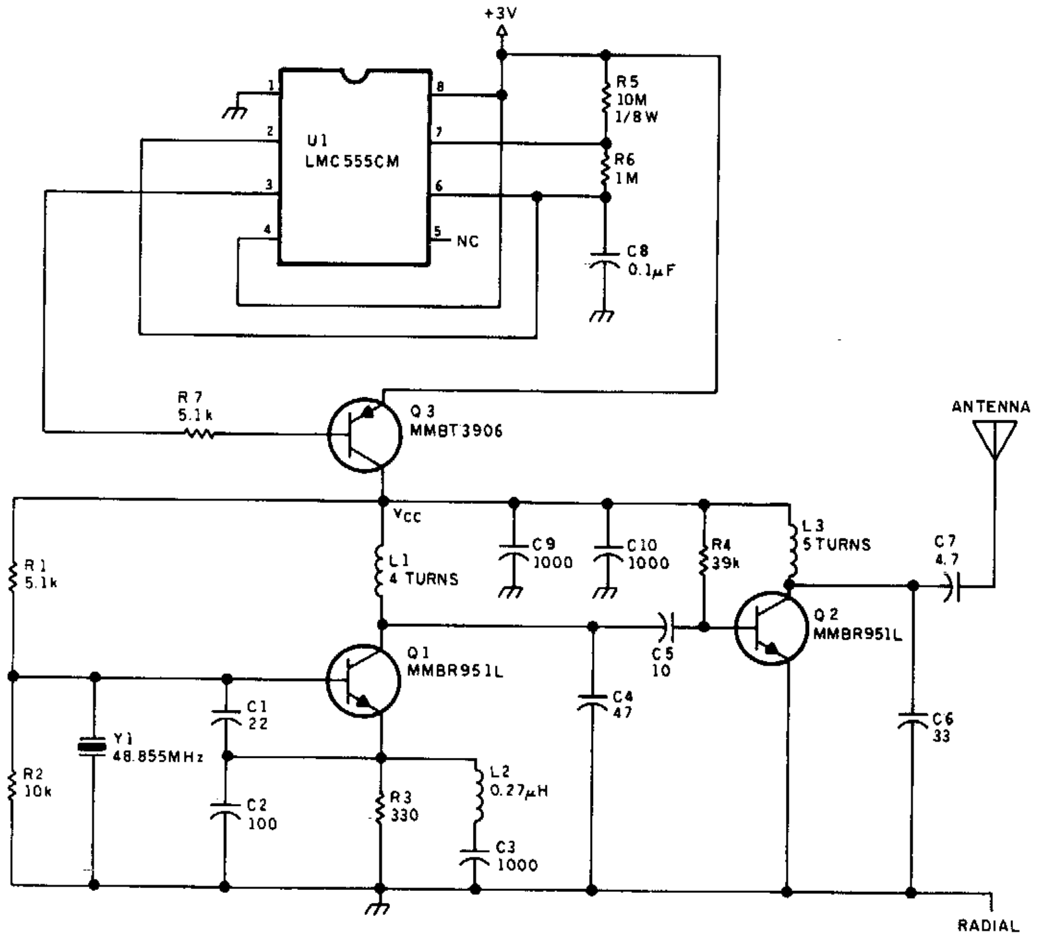
FM BUG



Mike	Electret condenser microphone
S1	SPST switch

McGRAW-HILL *Fig. 96-5*

LOW-POWER VHF BEACON TRANSMITTER

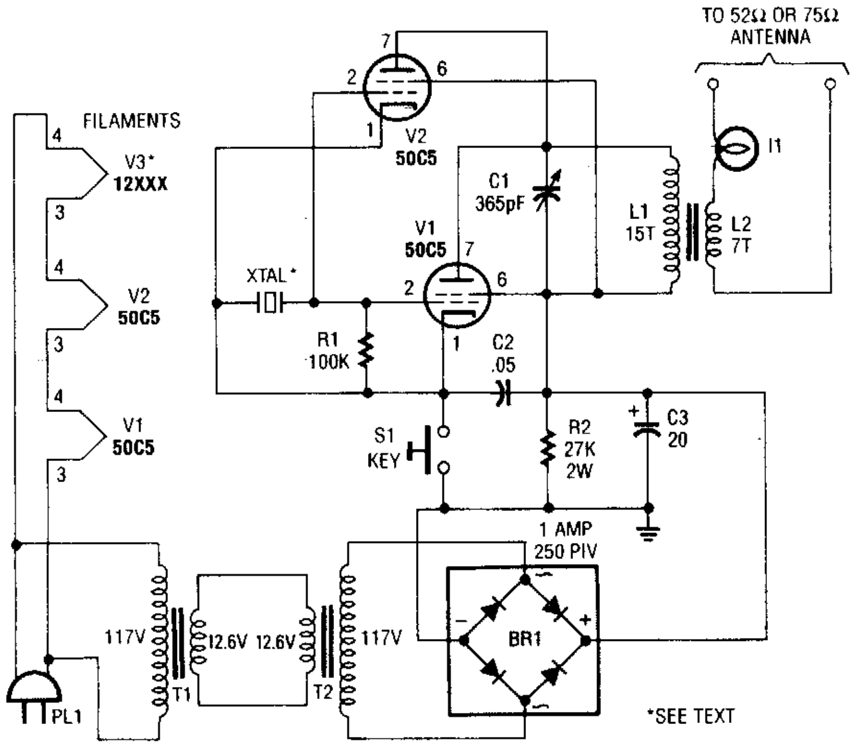


73 AMATEUR RADIO TODAY

Fig. 96-6

A crystal oscillator and tripler make up the low power beacon transmitter. U1 generates a pulse that keys the transmitter at a 10:1 duty cycle (100 ms on, 1 s off) to conserve battery power. This transmitter was used as a locator beacon.

LOW-COST 6-W, 40-M CW TRANSMITTER



POPULAR ELECTRONICS *Fig. 96-7*

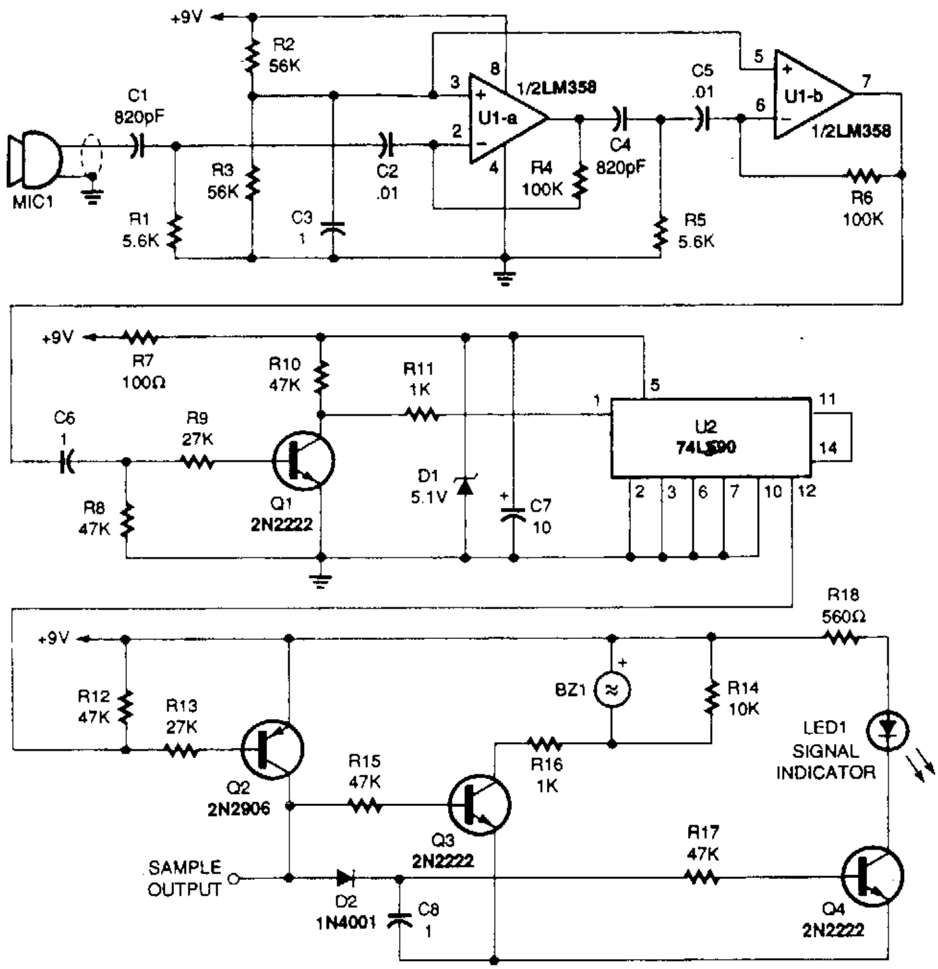
L1 and L2 are wound on a ⅞" diameter form. L1 is 15 turns #22 plastic-covered wire, and L2 is 7 turns #22 plastic-covered wire.

97

Ultrasonic Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ULTRASONIC REMOTE-CONTROL TESTER

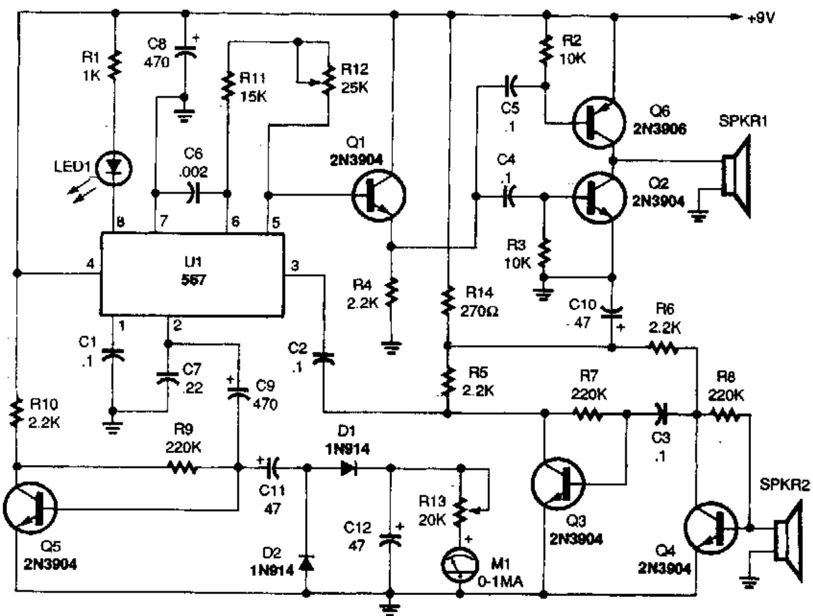


POPULAR ELECTRONICS

Fig. 97-1

This circuit picks up the ultrasonic tone via MIC1, amplifies it, and divides it by 10 in IC U2, a 74LS90. The output of U2 drives an audio amplifier and a piezoelectric element is used as a speaker.

ULTRASONIC MOTION DETECTOR



POPULAR ELECTRONICS *Fig. 97-2*

A 567 PLL IC operates in a dual-function mode as a signal generator and an FM receiver. The 567's square-wave output at pin 5 is coupled to the base of Q1, and from Q1's emitter to the input of the power amplifier (Q2 and Q6). The output drives the piezo speaker, SPKR1.

The receive portion of the circuit operates as follows: transistors Q3 and Q4 are connected in a two-stage, high-gain, audio-frequency amplifier circuit, with the input connected to a second piezo speaker (SPKR2) operating as a sensitive microphone. The amplifier's output is coupled to the 567's input at pin 3. When an in-band signal is received, the LED lights.

The 567's FM output is coupled from pin 2 to the input of a very-low-frequency single-transistor amplifier, Q5. The amplifier's output at Q5's collector drives a voltage-doubler circuit (C11, D1, D2, and C12). The dc output feeds a 0- to 1-mA analog meter.

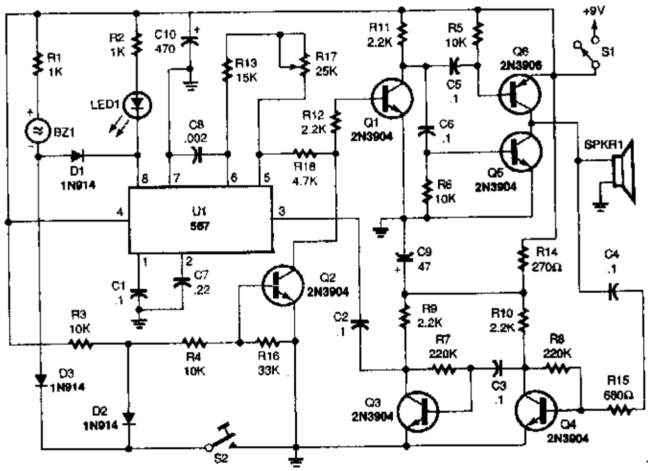
By placing the two piezo speakers one foot apart and aiming them in the same direction, toward a nonmoving solid object, the signal from the transmitter's speaker will reflect back into the receiver's speaker, and the frequency at the 567's input will be the same as the one being transmitted.

The ac output at pin 2 is zero when the outgoing and incoming frequencies are the same. However, when the signal is reflected from a moving object, the received frequency will be either lower or higher than the transmitted one. If the object is moving away from the speakers, the received frequency will be lower; if the object is moving toward the speakers, the frequency will be higher.

The pin-2 signal is fed through a 470-µF capacitor to the base of Q5, where the signal is amplified and fed to a voltage doubler, and then on to a meter circuit.

If you wish, the voltage doubler and meter can be removed and replaced with headphones connected between the negative side of C11 and circuit ground. That will allow you to listen to the difference-frequency signal as objects move in front of the speakers.

ULTRASONIC CW TRANSCEIVER



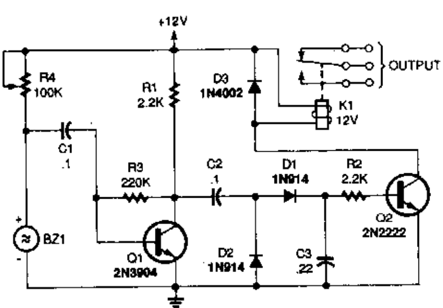
POPULAR ELECTRONICS **Fig. 97-3**

With the telegraph key at S2 "up" (open), the 567 PLL's input at pin 3 is coupled to the output of Q3. Transistors Q3 and Q4 are operating in a high-gain, two-stage audio-amplifier circuit. The piezo speaker is coupled to the input of the amplifier though a 0.1-µF capacitor and a 680-Ω isolation resistor.

In the receive mode, the piezo speaker operates as a sensitive microphone. Ultrasonic signals travel from the microphone through the two-stage amplifier to the input of the 567, and, if the signal's frequency is within the IC's bandwidth, the LED will light and piezo-sounder BZ1 will sing out for each "dit" and "dah" received. The receiver can be tuned to the incoming ultrasonic signal by adjusting R17. Of course, adjusting that potentiometer also changes the transmitter's frequency.

The transmitter operates each time the S2 is closed. When the key is closed, diode D3 supplies a path to ground for BZ1, causing that sounder to produce an audible signal for each "dit" and "dah" transmitted. Also, Q2's bias is taken to ground through D2, allowing Q1 to pass the 567's square-wave signal on to the input of the power amplifier and out through the speaker.

ULTRASONIC PROXIMITY SENSOR

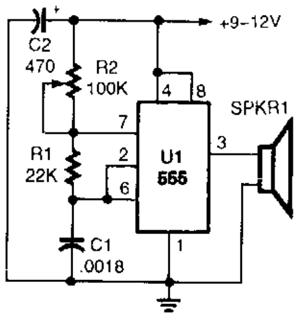


POPULAR ELECTRONICS **Fig. 97-4**

A 100-kΩ potentiometer, R4, sets the current fed to the Sonalert sounder. The potentiometer is adjusted to a point where the sounder just begins to make an audible sound. A single-transistor audio amplifier (Q1) is coupled to the positive side of the sounder and its output is fed to a voltage-doubler circuit. The doubler's dc output drives the base of Q2, which, in turn, operates the relay (K1). As long as the Sonalert is producing a sound, the relay stays energized.

When a solid object is moved in close proximity to the front of the sounder, the *Q* of the piezo element is lowered and the Sonalert's internal circuit ceases to operate; as a result, the relay drops out. By carefully adjusting R4, the circuit can be made quite sensitive.

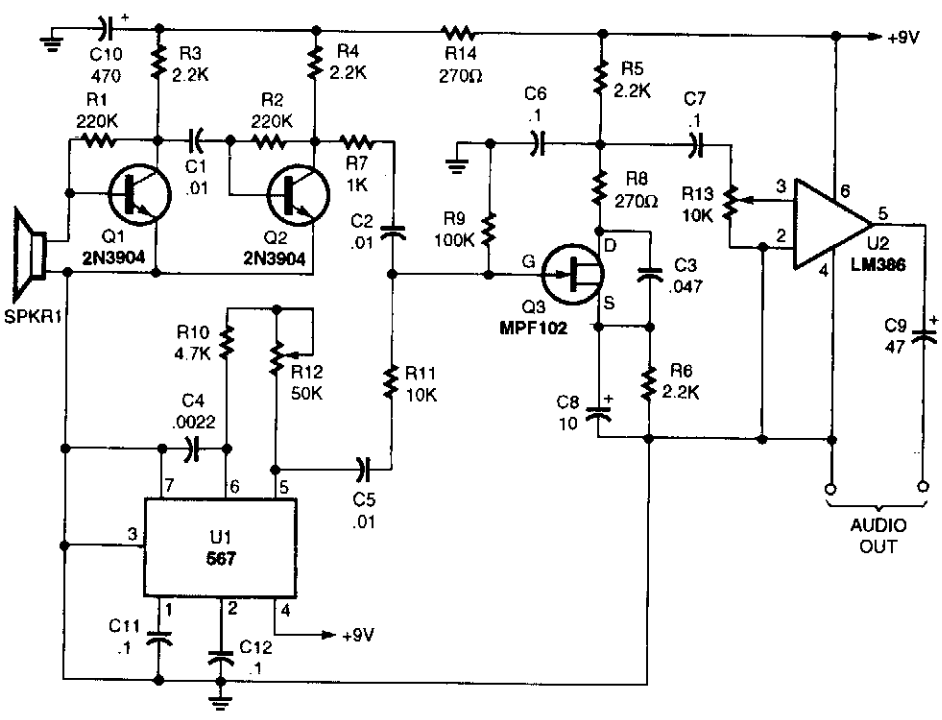
SIMPLE ULTRASONIC GENERATOR



This basic ultrasonic generator can be built easily and quickly. An NE555 drives a speaker. The frequency range is 12 to 50 kHz. SPKR1 is a piezo tweeter, etc.

POPULAR ELECTRONICS ***Fig. 97-5***

ULTRASONIC SOUND RECEIVER



POPULAR ELECTRONICS ***Fig. 97-6***

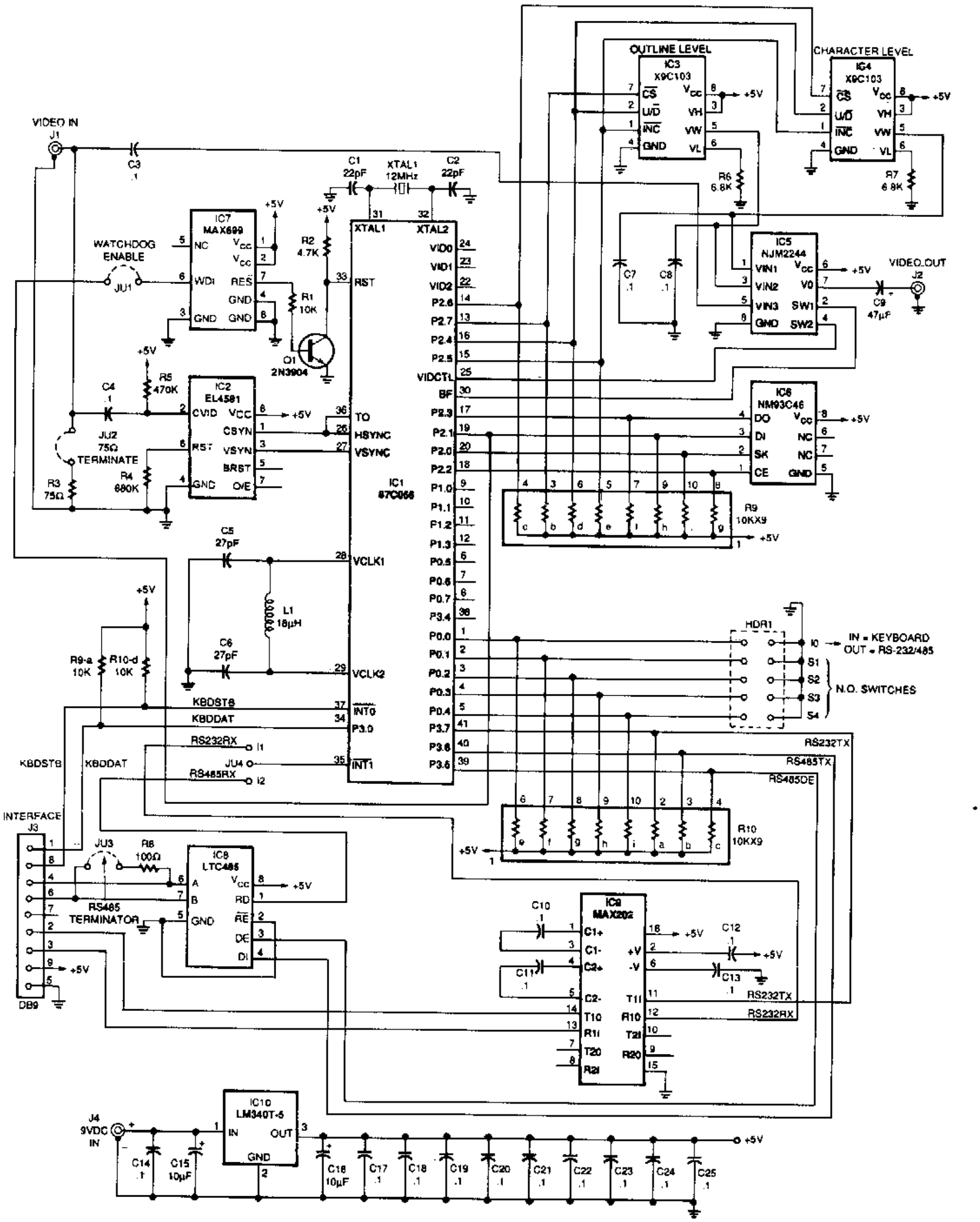
You won't be disappointed with the performance of this sensitive ultrasonic receiver. It can let you listen to bugs, bats, engines, and virtually any other source of ultrasonic sounds. This circuit uses a piezo tweeter as an ultrasonic microphone, amplifier stages Q1, Q2, and an LO using a 567 IC. Q3 is a mixer that heterodynes the ultrasonic sounds down to the audible range. U2 is an amplifier that will drive a pair of headphones.

98

Video Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

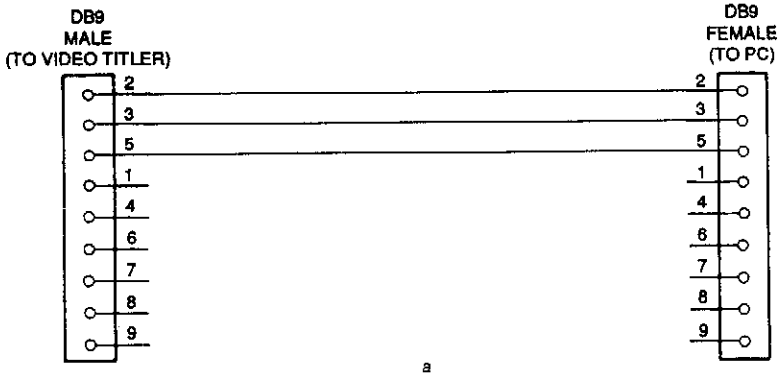
VIDEO TITLER



ELECTRONICS NOW!

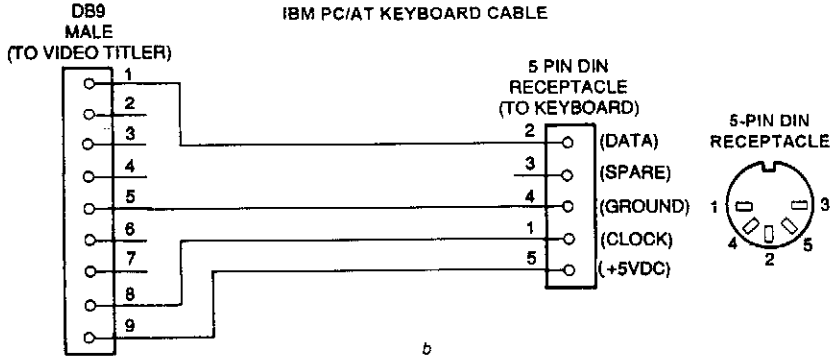
Fig. 98-1

IBM PC/AT RS232 SERIAL CABLE



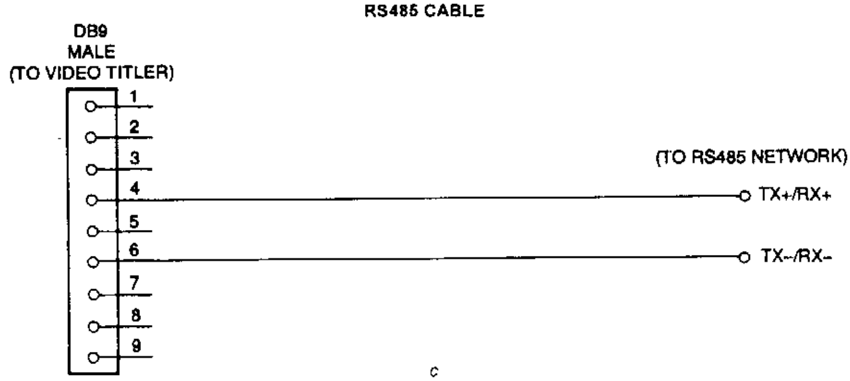
a

IBM PC/AT KEYBOARD CABLE



b

RS485 CABLE



c

There is a different cable for each interface. The RS-232 serial cable is shown in a, the PC/AT keyboard cable is shown in b, and the RS-485 cable is shown in c.

The figure shows the schematic of the video titler circuit. The power-on reset function is generated by IC7, a Maxim MAX699 reset, and watchdog pulse generator. That device supplies a reset pulse of 140 to 500 ms at power-up. This is accomplished with some external parts, as well as the OSD controller in IC1. First, the horizontal and vertical sync from the composite video input is detected by IC2, which is set for NTSC specification horizontal and vertical synchronization timing via resistor R4.

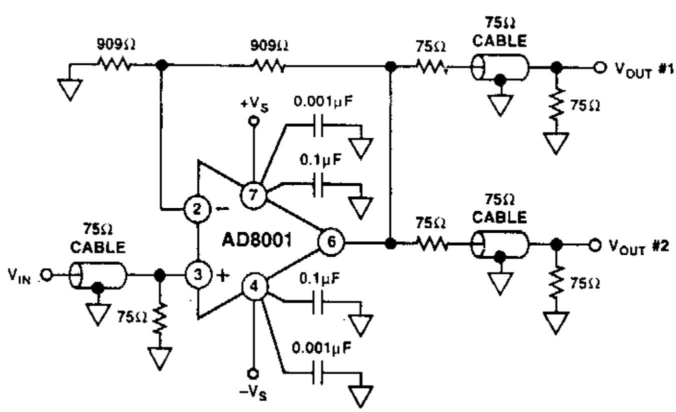
The detected horizontal and vertical sync is fed to IC1. The OSC controller in IC1 uses these signals to internally synchronize the overlay text to the incoming video. The frequency of the dot clock is controlled by components L1, C5, and C6. Text is overlaid by video multiplexer IC5, which is controlled by IC1.

The overlay character outline and intensity are controlled via solid-state potentiometers, allowing the microcontroller to control the position of their wipers and store the settings in an onboard EEPROM. The microcontroller's OSC logic controls the multiplexer timing from the BF (IC1, pin 30) and VIDCTL (IC1, pin 25) signals. The BF signal switches the video multiplexer between character and character-outline video, and VIDCTL switches the multiplexer between the input video and the overlay video from IC1. The dc levels from IC3 and IC4 set the character and outline intensity, and these levels are fed to video multiplexer IC5.

The video titler can store and recall text from EEPROM IC6, which has enough capacity to store one overlay screen and other required data, such as network address, horizontal and vertical overlay fine position, and type of interface.

The RS-232 interface is provided by a MAX202 transceiver. The RS-485 interface is provided by an LTC485 transceiver that provides both transmit and receive functions. The keyboard interface is basically a direct connection to the microprocessor.

VIDEO AMPLIFIER

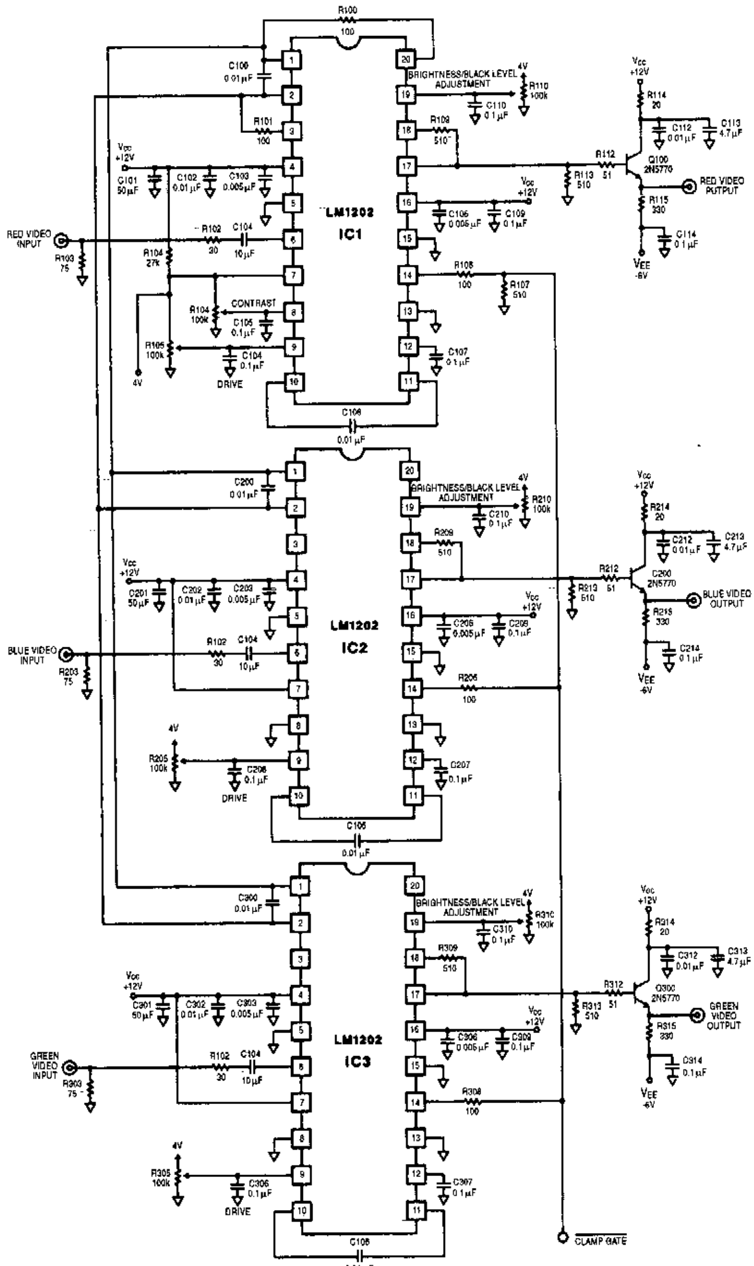


ANALOG DEVICES

Fig. 98-2

The AD8001 has been designed to offer outstanding performance as a video line driver. The important specification of differential gain (0.01%) and differential phase (0.025°) meet the most exacting HDTV demands for driving one video load. The AD8001 also drives up to two back-terminated loads with equally impressive performance (0.01%, 0.07°). Another important consideration is isolation between loads in a multiple-load application. The AD8001 has more than 40 dB of isolation at 5 MHz when driving two 75-Ω terminated loads.

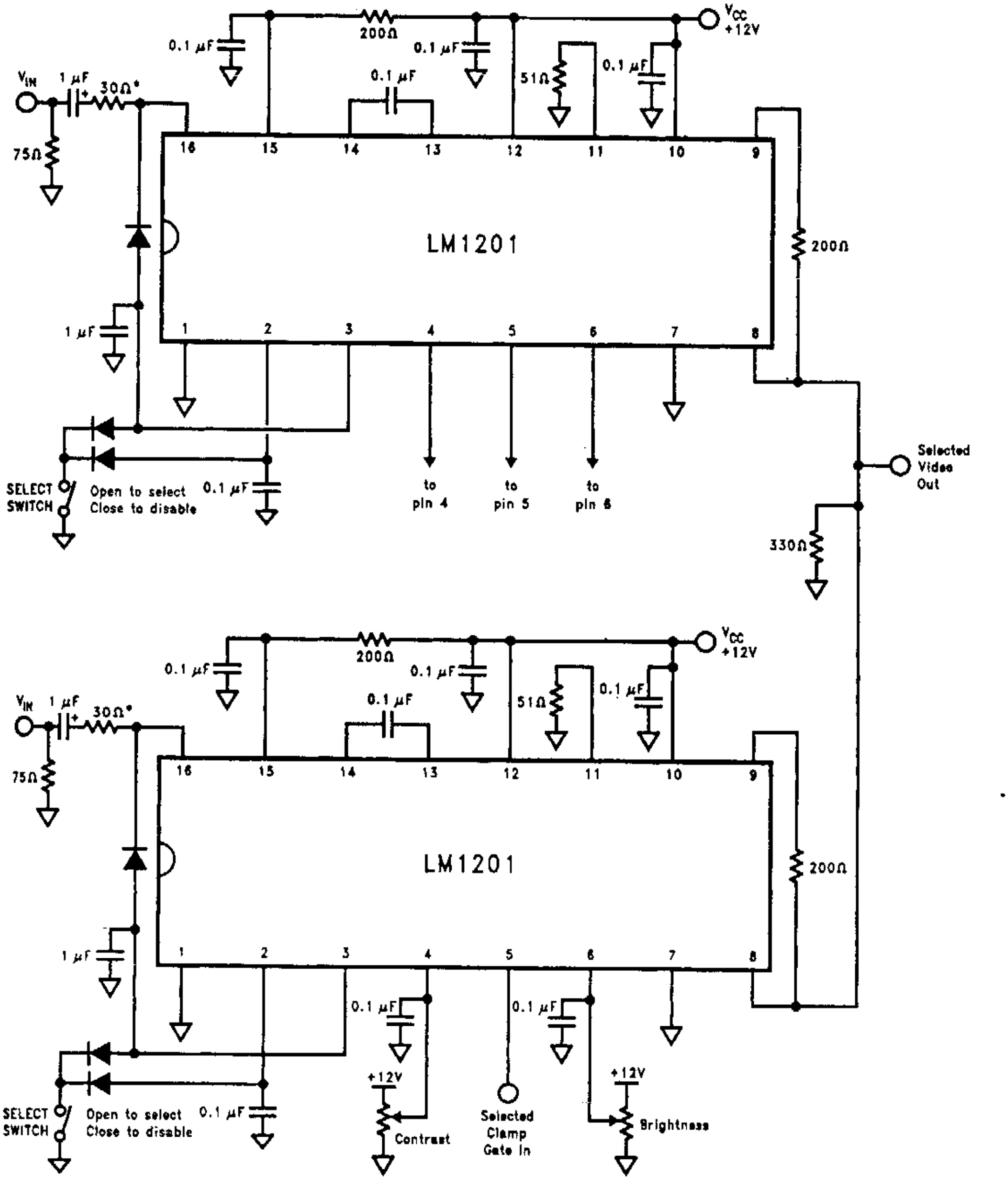
RGB VIDEO AMPLIFIER



NATIONAL SEMICONDUCTOR *Fig. 98-3*

This circuit is a three-channel RGB video amplifier with individual brightness, black level and drive controls.

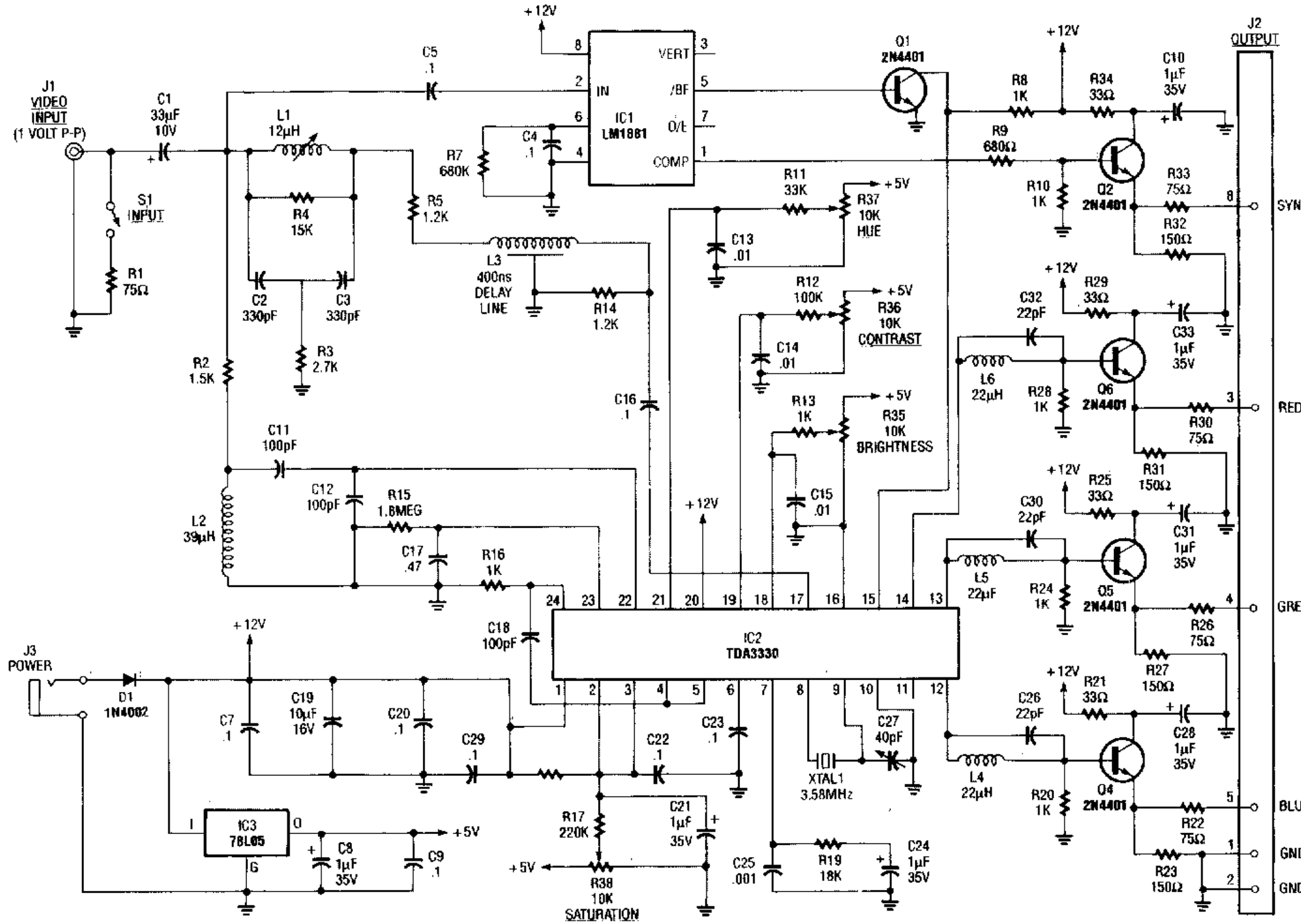
ONE-OF-TWO VIDEO SELECTOR



NATIONAL SEMICONDUCTOR

Fig. 98-4

NTSC-TO-RGB CONVERTER

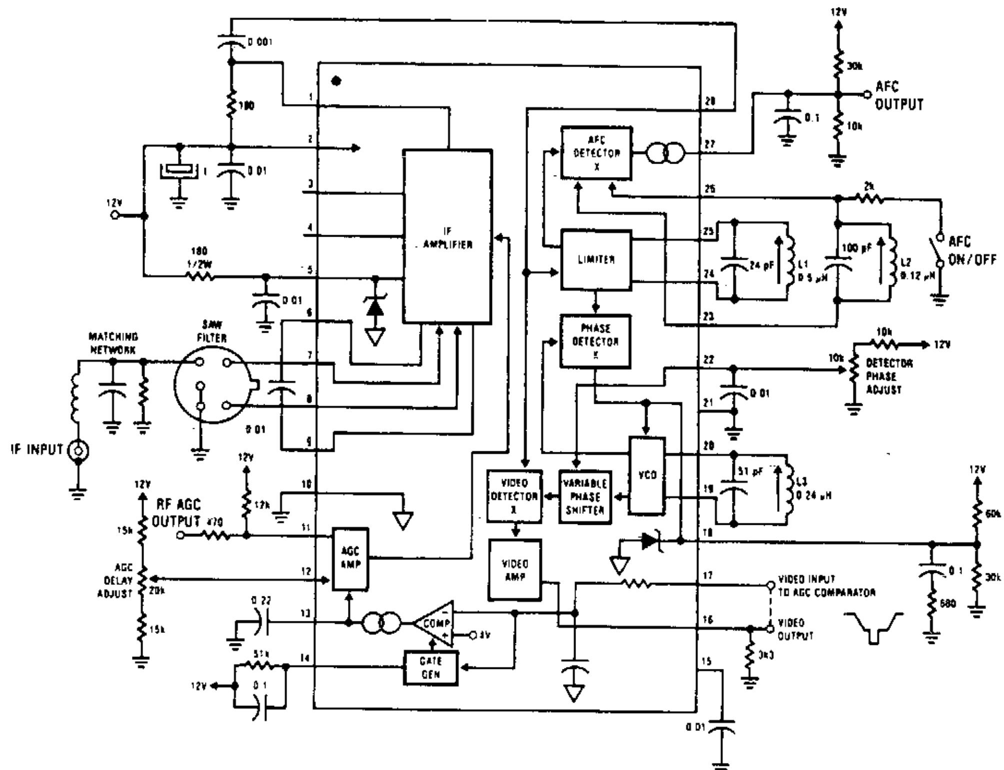


RADIO-ELECTRONICS

Fig. 98-5

This circuit takes baseband NTSC video, decodes it, and derives RGB video suitable for driving a color multisync computer monitor. This enables the user to take advantage of the generally better resolution of computer monitors.

VIDEO IF AMPLIFIER/DETECTOR



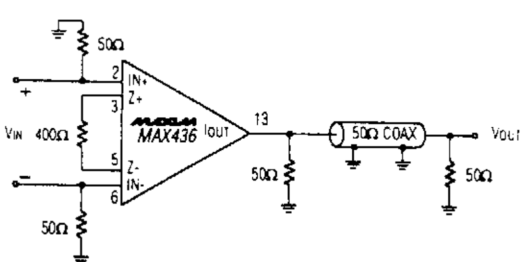
SAW Filter-MuRata SAF45MC/MA

L1-9 ½T ⎫ #22 wire
L2-4 ½T ⎬ on 3.16" form with
L3-6 ½T ⎭ HF core, shielded

All caps in uF unless noted

NATIONAL SEMICONDUCTOR *Fig. 98-6*

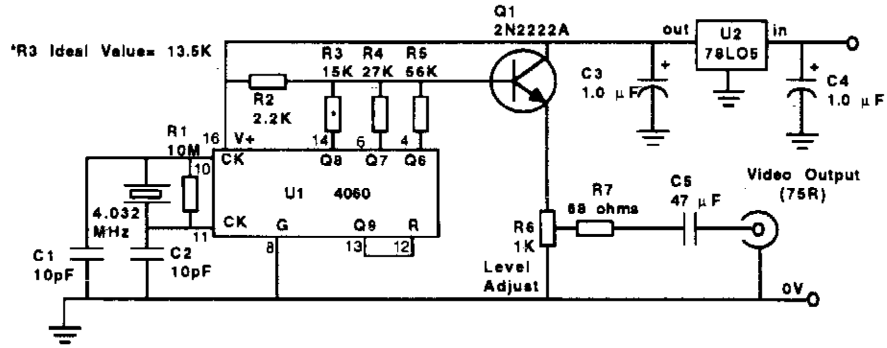
VIDEO CABLE DRIVER



MAXIM *Fig. 98-7*

This is a MAX436 coaxial-cable driving circuit.

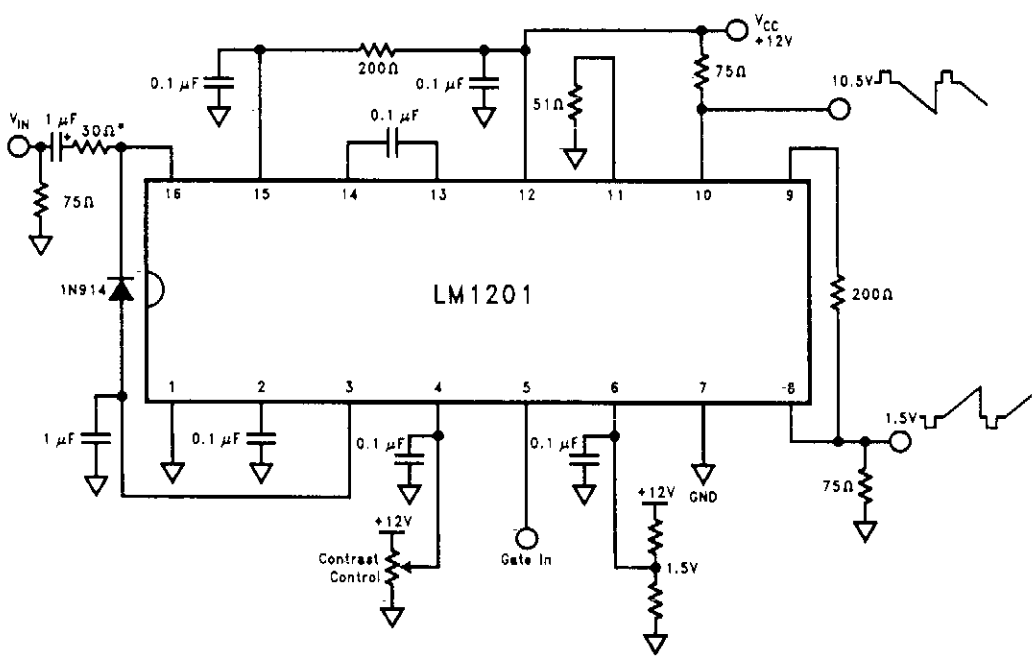
SIMPLE NTSC GRAY-SCALE VIDEO GENERATOR



73 AMATEUR RADIO TODAY **Fig. 98-8**

A 4.032-MHz crystal oscillator (256 × horizontal line scan rate) drives a BCD counter. The binary outputs of the counter are fed to R2 through R5, a simple weighting network for D/A conversion, resulting in a staircase video output with a rep rate of 15.75 kHz. This circuit should be useful for amateur TV linearity testing and setup purposes.

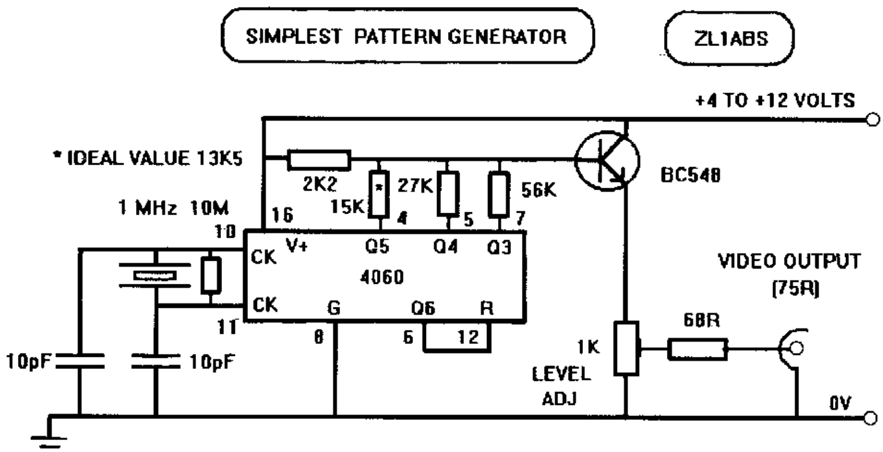
LM1201 VIDEO AMPLIFIER



NATIONAL SEMICONDUCTOR **Fig. 98-9**

This video amplifier has 75-Ω bi-phase outputs.

SIMPLE VIDEO GRAY-SCALE GENERATOR EUROPEAN LINE STANDARD

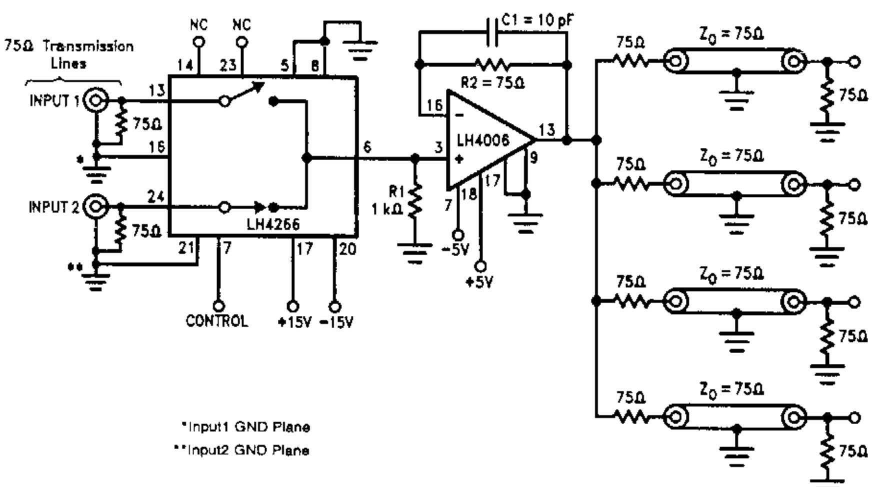


73 AMATEUR RADIO TODAY

Fig. 98-10

A simple gray-scale generator (staircase waveform) can be obtained with a CD4060 counter, a 1-MHz crystal oscillator, and several resistors to act as an elementary D/A converter to convert the binary count output to analog equivalent. This circuit is for European (PAL) standards.

VIDEO SWITCH

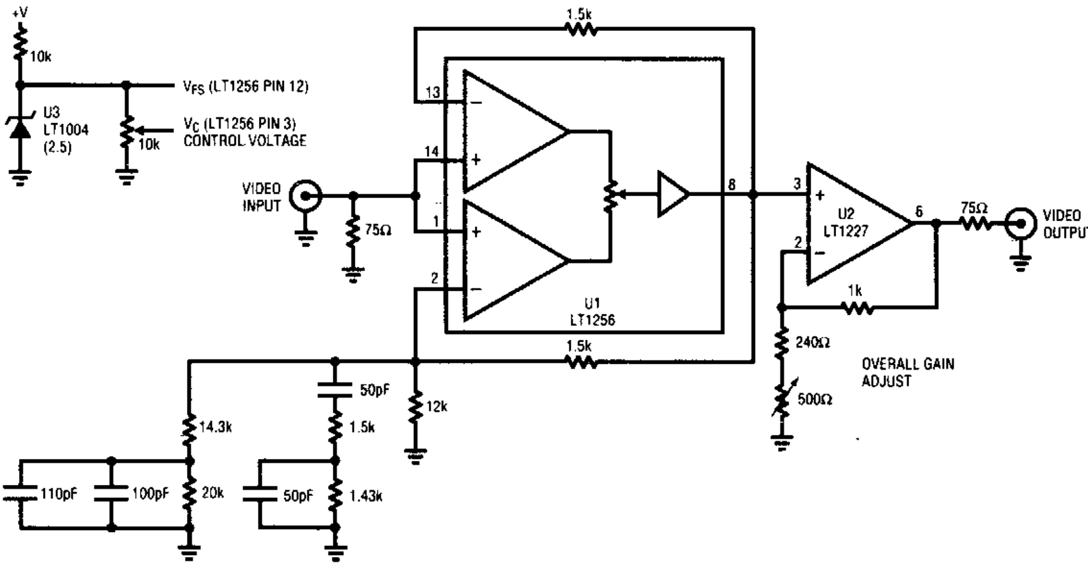


NATIONAL SEMICONDUCTOR

Fig. 98-11

Using National Semiconductor LH4266 and LH4006, this circuit switches one of two inputs to four output (75 Ω) lines.

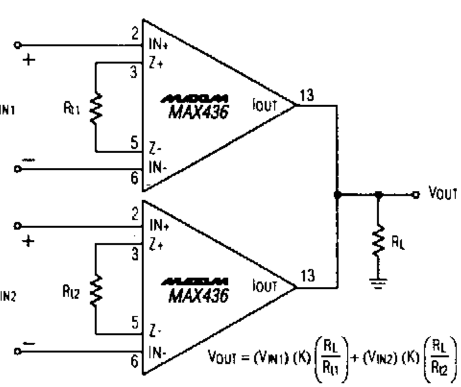
ADJUSTABLE VIDEO-CABLE EQUALIZER



LINEAR TECHNOLOGY *Fig. 98-12*

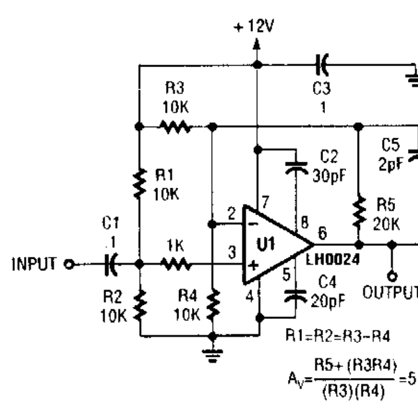
The figure is a complete schematic of the cable equalizer. The LT1256 (U1) is a two-input/one-output 40-MHz current feedback amplifier with a linear control circuit that sets the amount that each input contributes to the output. One amplifier (input pins 13 and 14) of the LT1256 is configured as a gain of one with no frequency equalization. The other amplifier (input pins 1 and 2) has frequency equalizing components in parallel with the 12-kΩ gain resistor. An additional amplifier (U2, LT1227) is used to set the overall gain. Two amplifiers were used here to make setting the gain a single adjustment, but in a production circuit, the LT1256 can be configured to have the necessary gain and the whole function can be done with one chip.

VIDEO SUMMING AMPLIFIER



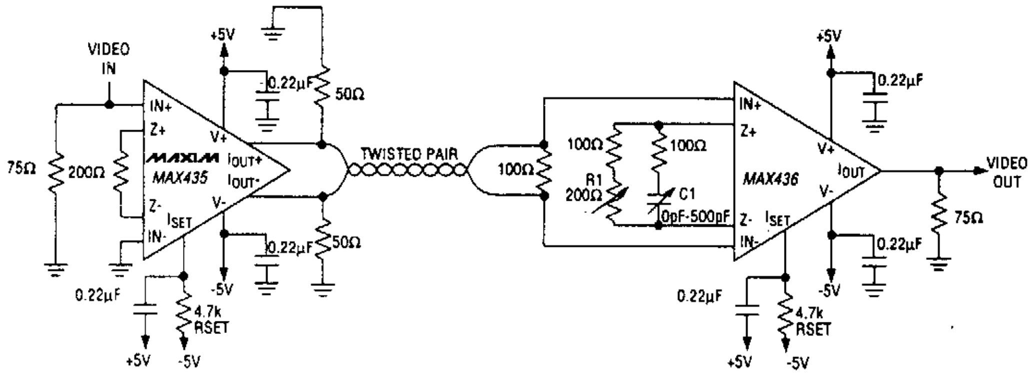
MAXIM *Fig. 98-13*

VIDEO AMPLIFIER



POPULAR ELECTRONICS *Fig. 98-14*

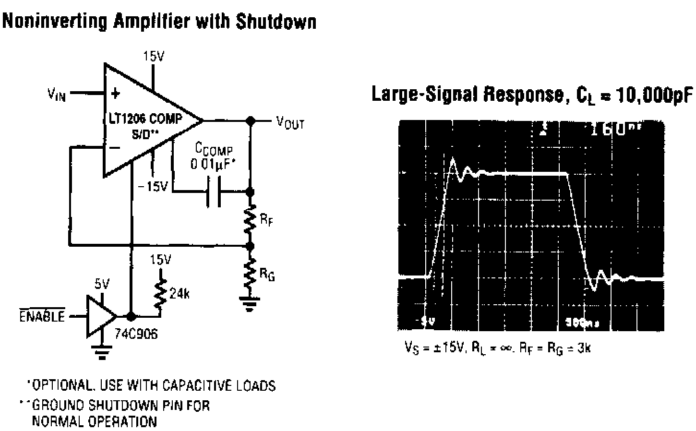
TWISTED-PAIR VIDEO DRIVER/RECEIVER CIRCUIT



MAXIM *Fig. 98-15*

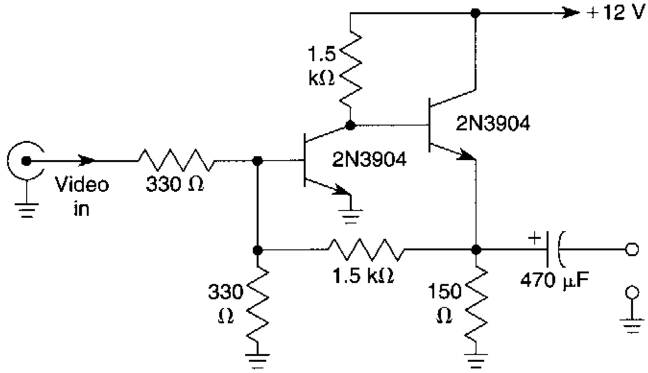
This circuit should be useful where a twisted-pair video line is to be used. R1 is adjusted for proper gain (monitor brightness and contrast) and C1 is adjusted for best color.

250-mA 60-MHz CURRENT-FEEDBACK AMPLIFIER FOR VIDEO APPLICATIONS



LINEAR TECHNOLOGY *Fig. 98-16*

VIDEO DRIVER/AMPLIFIER

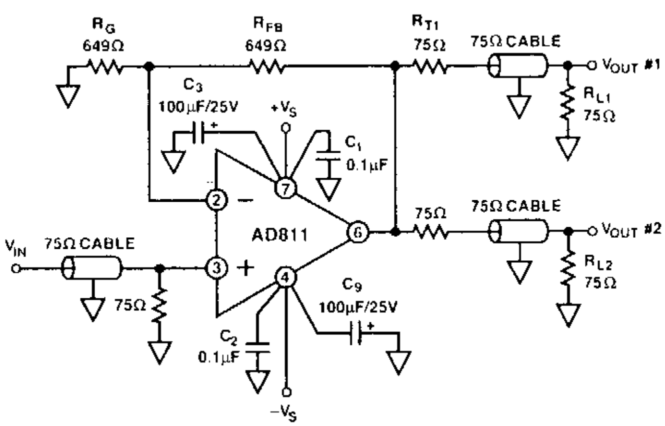


WILLIAM SHEETS

Fig. 98-17

This simple circuit has a voltage gain of about 5× and will drive low-impedance loads (75 Ω) to 1.5 V p-p or better.

VIDEO LINE DRIVER



ANALOG DEVICES

Fig. 98-18

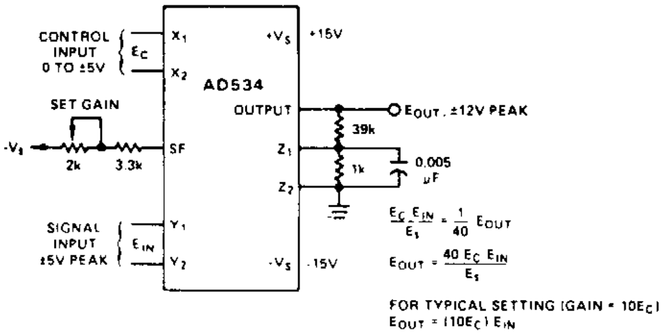
This video buffer/line driver operates at a gain of +2 and drives a pair of 75-Ω lines with 75-Ω back terminations. The overall terminated gain is unity.

99

Voltage-Controlled Amplifier Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

VOLTAGE-CONTROLLED AMPLIFIER

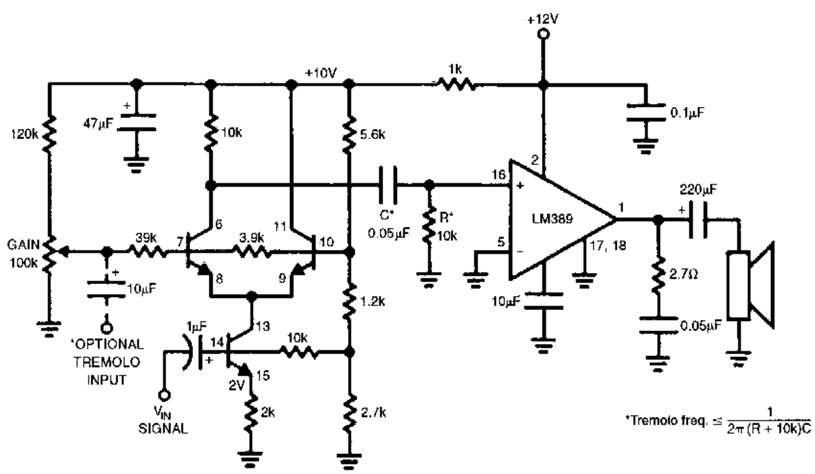


ANALOG DEVICES *Fig. 99-1*

A constant or varying signal applied to the X input, E_c, controls the gain for a constant or variable signal applied to the Y input, E_{in}. The inputs could be interchanged.

For this circuit, the "set gain" potentiometer is typically adjusted to provide a calibration for gain of Z 10 per-V-of-E_c. The bandwidth is dc to 30 kHz, independent of the gain. The wideband noise (10 Hz to 30 kHz) is 3 mV rms, typically, corresponding to full-scale signal-to-noise of 70 dB. Noise, referred to the signal input (E_c = ±5 V) is 60 μV rms, typically.

VOLTAGE-CONTROLLED AUDIO AMPLIFIER



NATIONAL SEMICONDUCTOR *Fig. 99-2*

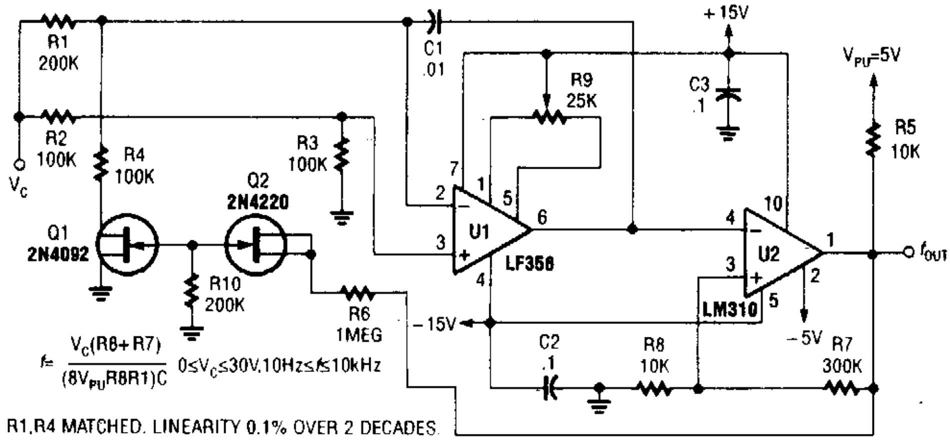
The LM389 has internal transistors used in this circuit.

100

Voltage-Controlled Oscillator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

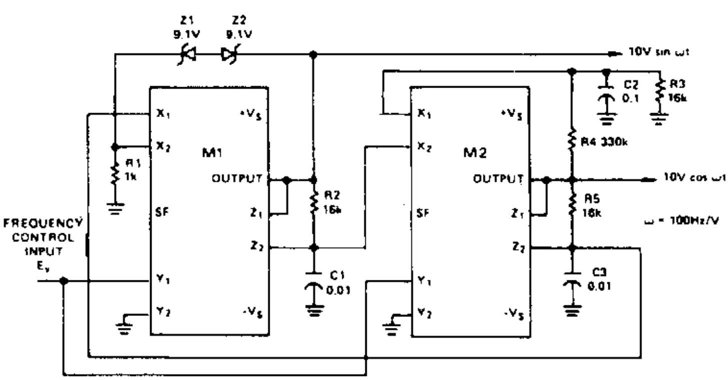
THREE-DECADE VCO



POPULAR ELECTRONICS *Fig. 100-1*

A range of 10 Hz to 10 kHz is covered by this circuit.

VOLTAGE-CONTROLLED TWO-PHASE OSCILLATOR



ANALOG DEVICES *Fig. 100-2*

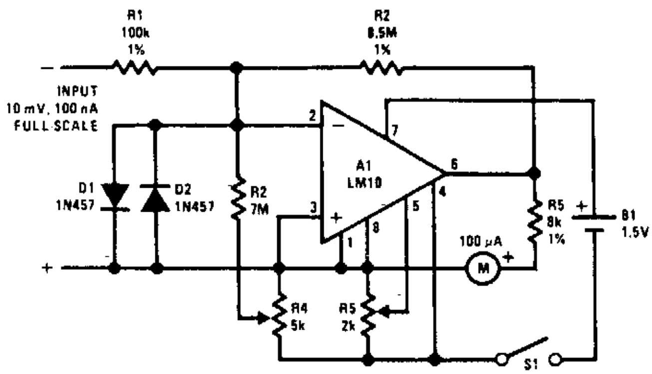
This circuit uses two multipliers for integration-with-controllable-time-constants in a feedback loop. R2 and R5 will be recognized in the AD534 voltage-to-current configuration; the currents are integrated in C1 and C3, and the voltages they develop are connected at high impedance in proper polarity to the X inputs of the "next" AD534. The frequency-control input, EY, varies the integrator gains, with a sensitivity of 100 Hz/V, and frequency error typically less than 0.1% of full scale from 0.1 V to 10 V (10 Hz to 1 kHz). C2 (proportional to C1 and C3), R3, R4 provide regenerative damping to start and maintain oscillation. Z_1 and Z_2 stabilize the amplitude at low distortion by degenerative damping above ±10 V.

101

Voltage-Measuring Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 706. The figure number in the box of each circuit correlates to the entry in the Sources section.

METER AMPLIFIER FOR 1.5-V SUPPLY

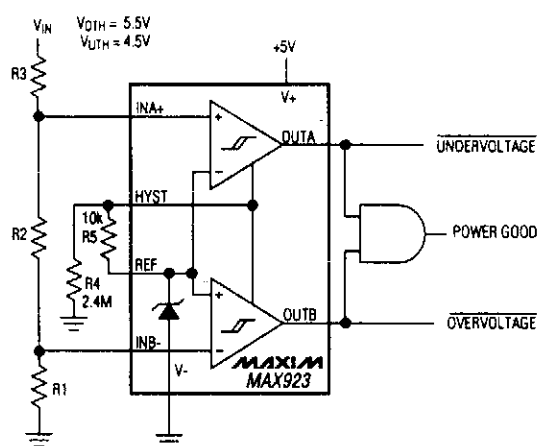


NATIONAL SEMICONDUCTOR

Fig. 101-1

An LM10 is used as a meter amplifier. Accuracy can be maintained over a 15°C to 55°C range for a full-scale sensitivity of 10 mV and 100 nA. The offset voltage error is nulled with R5, and the bias current can be balanced out with R4. The zeroing circuits operate from the reference output and are essentially unaffected by changes in battery voltage, so frequent adjustments should not be necessary. Total current drain is under 0.5 mA, giving an approximate life of 3 to 6 months with an "AA" cell and over a year with a "D" cell. With these lifetimes, an ON/OFF switch might be unnecessary. A test switch that converts to a battery-test mode might be of greater value.

VOLTAGE MONITOR

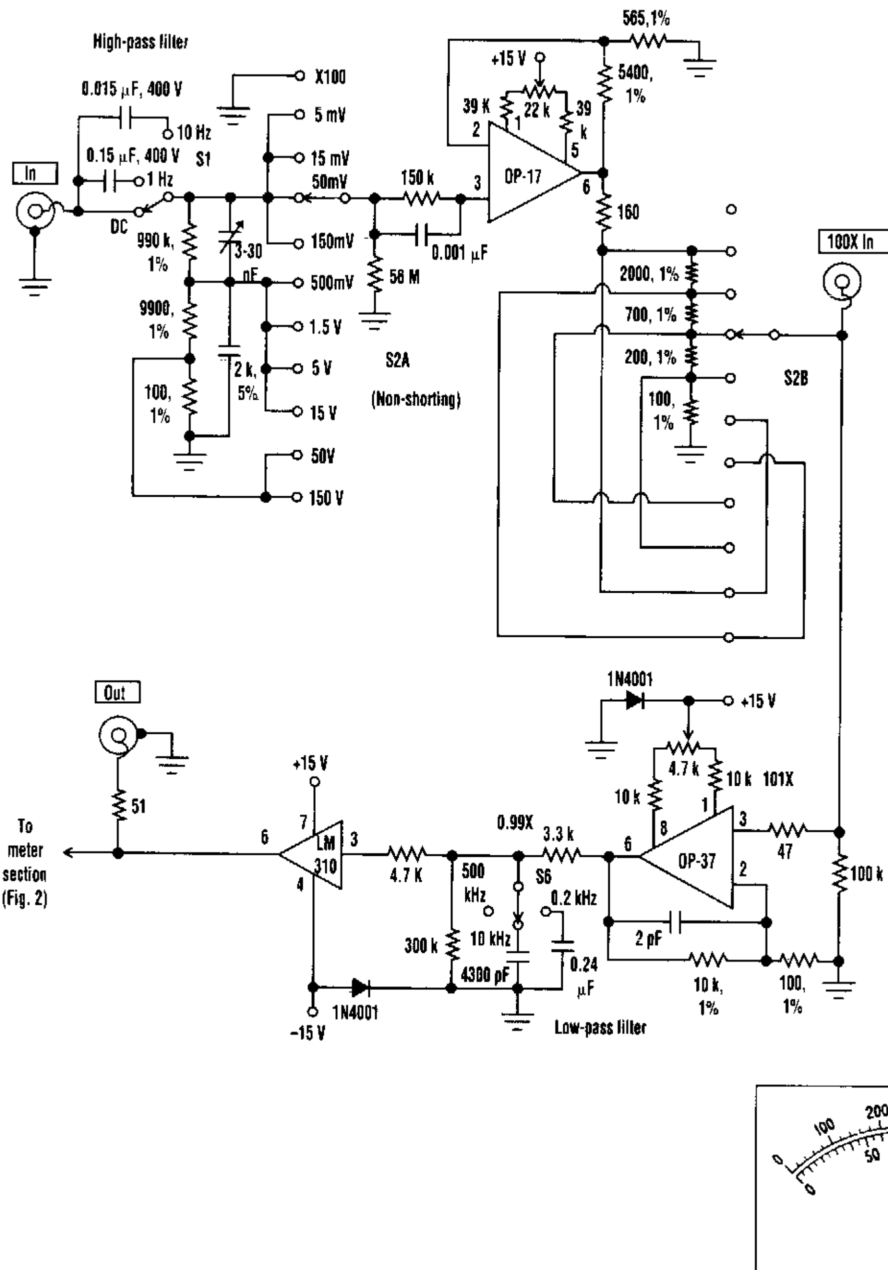


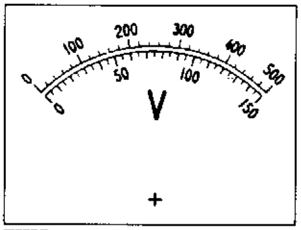
MAXIM

Fig. 101-2

A MAX923 dual comparator is used as a window detector. For a threshold of 4.5 V and 5.5 V, R_1 = 1.068 MΩ, R_2 = 61.9 kΩ, and R_3 = 1 MΩ.

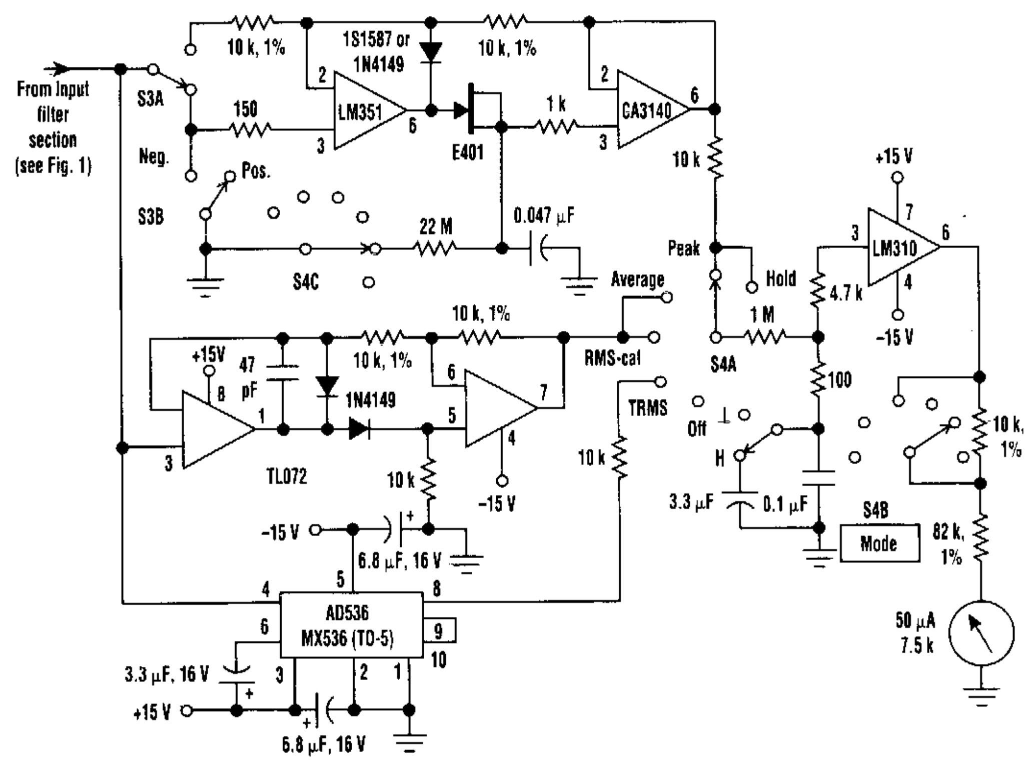
ac VOLTMETER HAS UNIQUE FEATURES





THIS meter scale can be enlarged on a copier and attached to a meter face for the dual-scale ac voltmeter.

Fig. 101-3



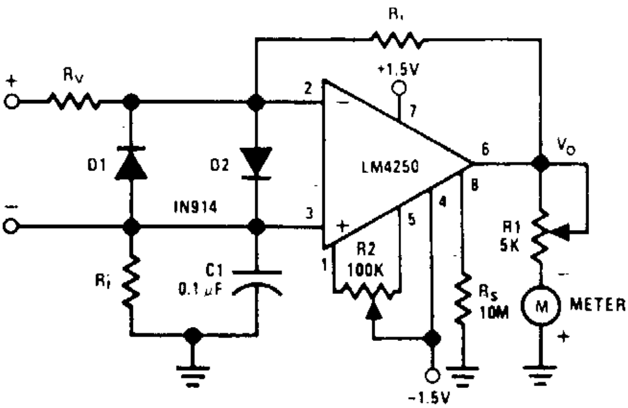
Though it's built with standard components, this ac voltmeter contains many features not typically found in commercial meters; the most unusual is a selection of rectification modes. The meter responses available include true RMS (TRMS), average, RMS-calibrated average responding, positive peak, negative peak, positive-peak hold, and negative-peak hold.

High- and low-pass filters (S1 and S6, respectively) allow the –3-dB-passband to be varied from as little as 10 Hz to 200 Hz, to as wide as dc to 500 kHz. The low-pass filter also is effective in the 100× amplifier mode, where the input equivalent noise level is only 0.3 µV, with 10-kHz roll-off.

dc VOLTMETER

Resistance Values for a DC Voltmeter

V FULL SCALE	R_V [Ω]	R_f [Ω]	R'_f [Ω]
10 mV	100k	1.5M	1.5M
100 mV	1M	1.5M	1.5M
1V	10M	1.M	1.5M
10V	10M	300k	0
100V	10M	30k	0

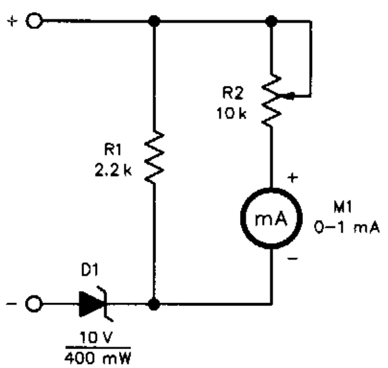


VOLTMETER

NATIONAL SEMICONDUCTOR ***Fig. 101-4***

A wide-range voltmeter circuit. This inverting amplifier has a gain varying from –30 for the 10-mV full-scale range to –0.003 for the 100-V full-scale range. Diodes D1 and D2 provide complete amplifier protection for input overvoltages as high as 500 V on the 10-mV range, but if overvoltages of this magnitude are expected under continuous operation, the power rating of R_v should be adjusted accordingly.

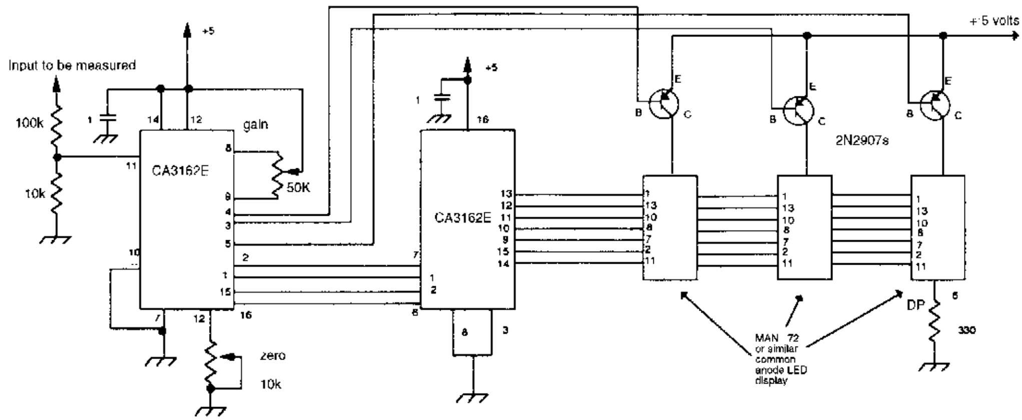
EXPANDED-SCALE dc METER FOR 12-V SYSTEMS



This circuit can be used to monitor a 12-V system with a meter reading 10 V to another voltage. Expect 10 to 20 V, depending on the setting of R2. Depending on the characteristics of D1, R1 might be increased or eliminated entirely.

QST ***Fig. 101-5***

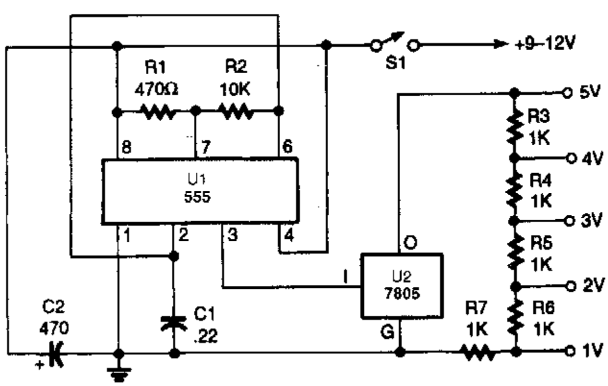
SIMPLE 3-DIGIT DVM



73 AMATEUR RADIO TODAY *Fig. 101-6*

A CA3162ZE A-D converter drives a CA3161 BCD decoder/driver and LED display to form a simple DVM circuit. The 50-kΩ gain control and 100-kΩ/10-kΩ voltage divider determine full-scale range.

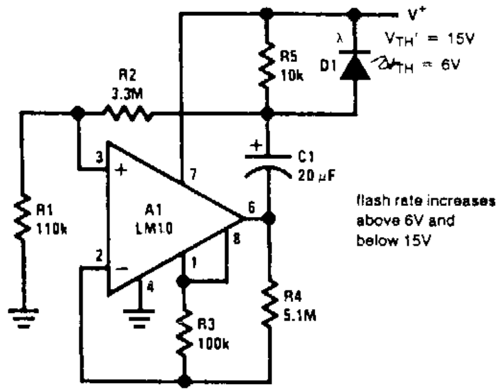
INEXPENSIVE VOLTAGE CALIBRATOR



POPULAR ELECTRONICS *Fig. 101-7*

In the voltage calibrator, two low-cost ICs—a 555 oscillator/timer and a 78055 5-V 1.5-A voltage regulator—along with a precision voltage divider network are used to provide outputs of 1- to 5-V peak-to-peak.

DOUBLE-ENDED VOLTAGE MONITOR



NATIONAL SEMICONDUCTOR

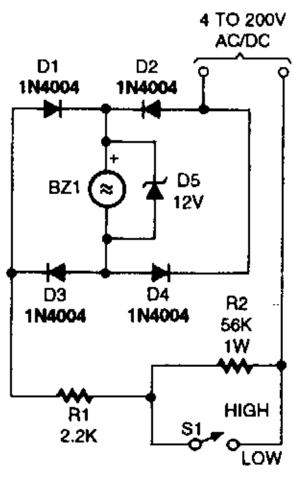
Fig. 101-8

This circuit has the added feature that it can sense an over-voltage condition. The lower activation threshold is given by equation (1), but above a threshold,

$$V_{TH} = \frac{R4(R_1 + R_2)\ V_{REF}}{R_1(R_3 + R_4) - R3(R_1 + R_2)}$$

oscillation again ceases. Below V_{TH}, the op amp output is saturated negative, but above V_{TH}, it is saturated positive. The flash rate approaches zero near either limit.

AUDIBLE VOLTAGE INDICATOR

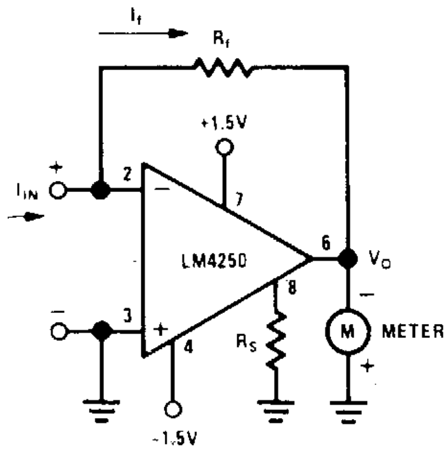


POPULAR ELECTRONICS

Fig. 101-9

The audible voltmeter can be used to test for ac or dc voltages in a circuit. With S1 closed, the circuit can be used to test for voltages between 4 and 24 V, and when S1 is open, it can be used to check for the presence voltages of up to 200 V.

LOW-DRAIN METER AMPLIFIER



NATIONAL SEMICONDUCTOR

Fig. 101-10

Meter amplifiers normally require one or two 9-V transistor batteries. Because of the heavy current drain on these supplies, the meters must be switched to the OFF position when not in use. The meter circuit described here operates on two 1.5-V flashlight batteries and has a quiescent power drain so low that no on/off switch is needed. A pair of Eveready No. 950 "D" cells will serve for a minimum of one year without replacement. As a dc ammeter, the circuit will provide current ranges as low as 100 nA full-scale.

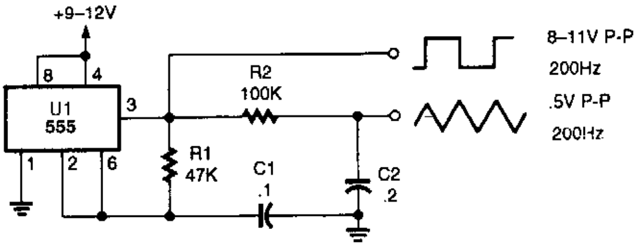
The basic meter amplifier circuit shown is a current-to-voltage converter. Negative feedback around the amplifier ensures that currents I_{IN} and I_f are always equal, and the high gain of the op amp ensures that the input voltage between pins 2 and 3 is in the microvolt region. Output voltage V_o is therefore equal to $-I_f R_f$. Considering the ±1.5-V sources (±1.2 V end of life) a practical value of V_o for full-scale meter deflection is 300 mV. With the master bias-current setting resistor (R_s) set at 10 MΩ, the total quiescent current drain of the circuit is 0.6 µA for a total power supply drain of 1.8 µW. The input bias current, required by the amplifier at this low level of quiescent current, is in the range of 600 pA.

102

Waveform Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE TRIANGLE WAVEFORM GENERATOR



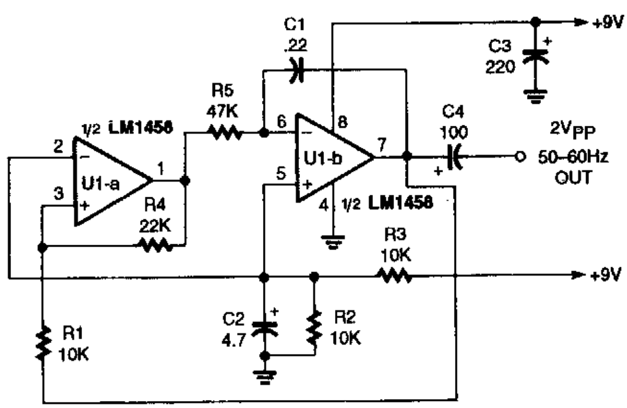
POPULAR ELECTRONICS

Fig. 102-1

The circuit is a triangle waveform-generator circuit that uses as few parts as possible. A 555 timer IC, two resistors, and two capacitors make the triangle waveform. The IC is connected in a 50% duty-cycle astable square-wave oscillator circuit. The square-wave output is fed from pin 3 of the IC to an RC shaping circuit.

When the 555's square-wave output goes high, C2 begins to charge through R2 and the voltage across C2 increases as long as the output remains high. When the IC's output goes low again, C2 begins to discharge through R2 reducing the voltage across C2 as long as the output remains low. The resulting waveform across C2 takes the shape of a triangle. The best waveform linearity is obtained when R_2 and C_2 are made as large as possible. With the component values shown, the peak-to-peak output is 0.5 V at a frequency of about 200 Hz.

TRIANGLE-WAVE GENERATOR

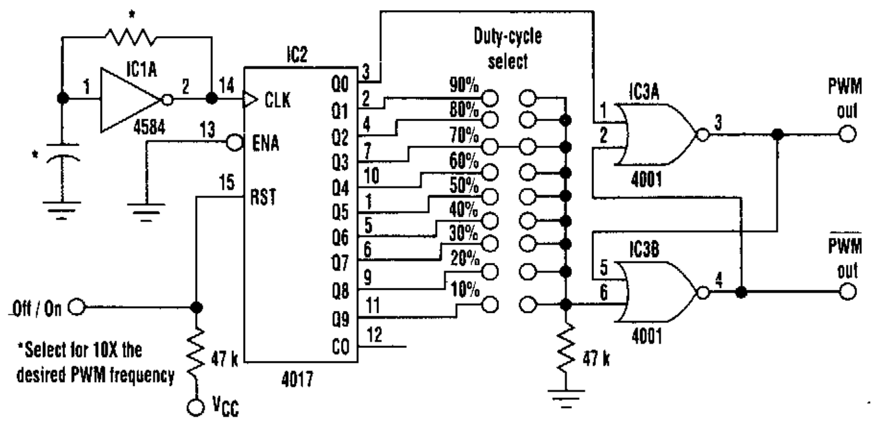


POPULAR ELECTRONICS

Fig. 102-2

This oscillator, which is built around an LM1458 dual op amp and a few inexpensive components, produces a 2-V peak-to-peak, triangular waveform.

GENERATE ACCURATE PWM SIGNALS



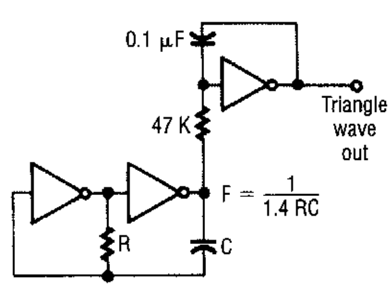
ELECTRONIC DESIGN *Fig. 102-3*

Accurate 10 to 90% duty-cycle PWM signals can be generated using this simple circuit setup. The desired duty cycle is selected by a single jumper block. PWM clock IC1 runs at 10× the desired pulse drive frequency. IC2, a 4017 divide-by-10 counter, decodes the clock pulses into one of 10 outputs. Output 0 resets IC3, the PWM latch. The latch stays reset until the desired duty-cycle output set by the jumper block is reached. At this point, the PWM latch is set, and the PWM output line remains high until output 0 is decoded again.

By calling IC2's output (0) the "reset" line for the latch, the PWM output is forced inactive if the jumper strap is removed to change duty cycles without first powering down.

Using the zero-state reset allows IC2's reset pin to be used as an on/off control line for the circuit. The complementary PWM output could be used in a full bridge design.

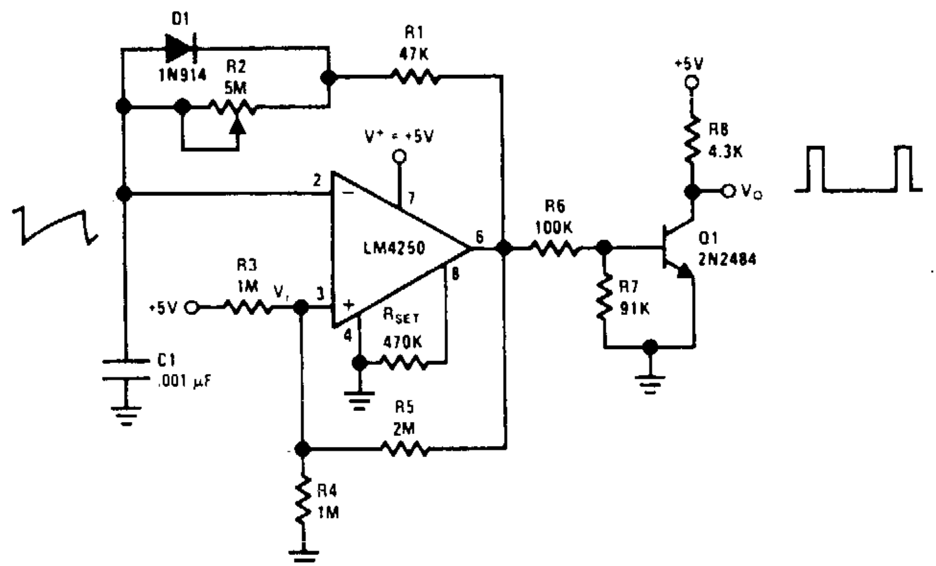
TRIANGLE-WAVE GENERATOR

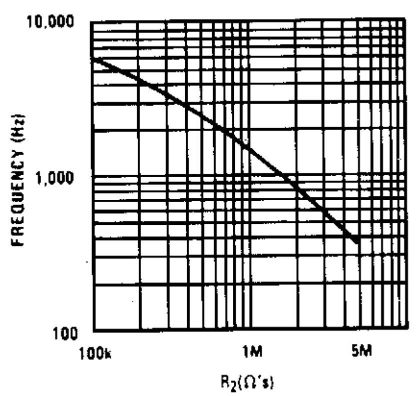


ELECTRONICS NOW *Fig. 102-4*

The first two gates are set up as a square-wave oscillator, and the last one makes the conversion to triangle waves.

LOW-FREQUENCY PULSE GENERATOR



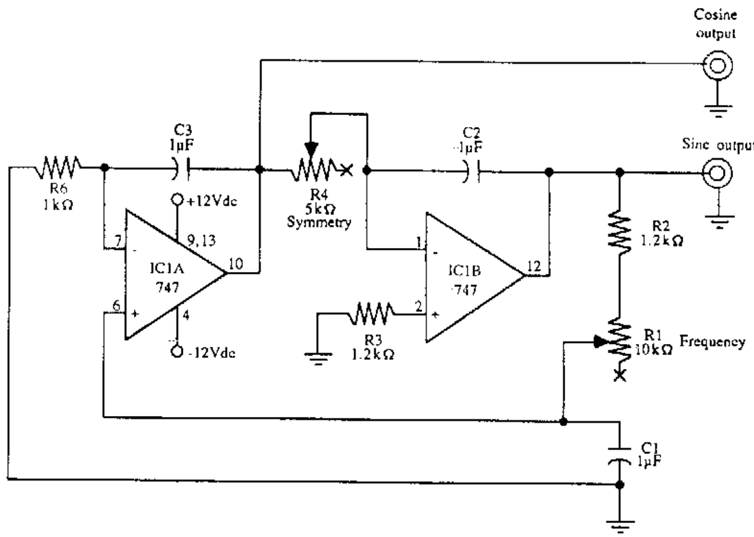


Pulse frequency vs. R_2

NATIONAL SEMICONDUCTOR

Fig. 102-5

SINE/COSINE AUDIO GENERATOR FOR GALVANOMETER EXPERIMENTS



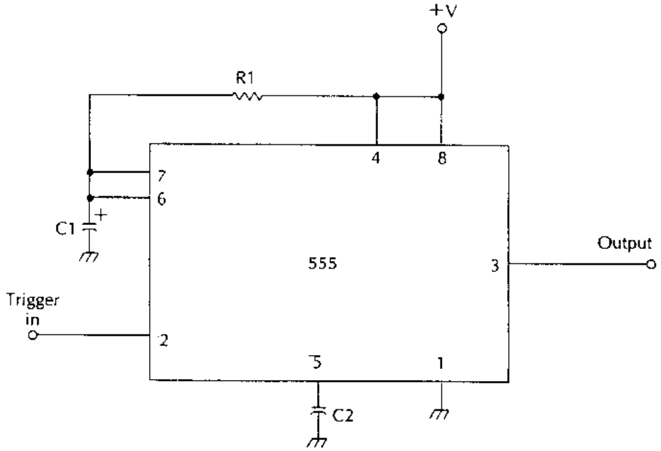
C1 – C3	1 μF polarized electrolytic
IC1	LM747 dual op amp IC
J1,J2	1/8-inch miniature phone jack

All resistors are 5 to 10 percent tolerance, 1/4 watt.
All capacitors are 10 to 20 percent tolerance, rated at 35 volts or more.

McGRAW-HILL **Fig. 102-6**

This circuit shows how to implement a sine/cosine audio generator for operating two galvanometers.

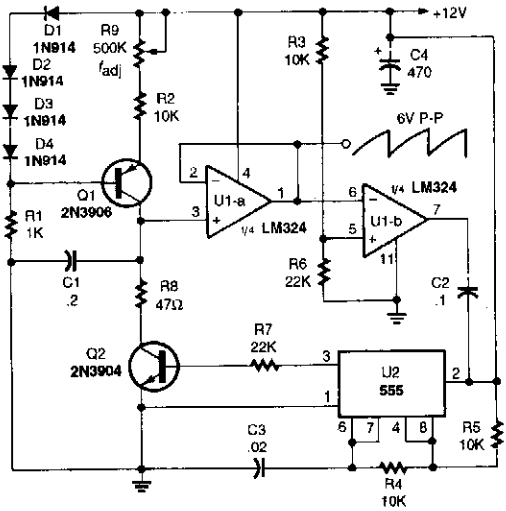
BASIC 555 MONOSTABLE



McGRAW-HILL **Fig. 102-7**

$$T + 1.1\,R_1C_1$$

OP-AMP SAWTOOTH GENERATOR

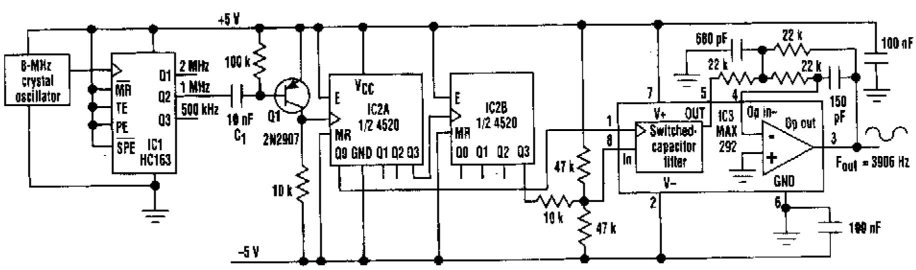


POPULAR ELECTRONICS *Fig. 102-8*

The sawtooth generator circuit shown is reset at the end of each cycle. The result is a constant peak-to-peak output throughout the circuit's frequency range.

The constant-current generator circuit, the voltage-follower circuit, and the comparator circuit produce the waveform. A 555 timer IC (U2) is configured as a one-shot multivibrator that's triggered by the comparator's negative output pulse.

DIGITAL SINE-WAVE GENERATOR



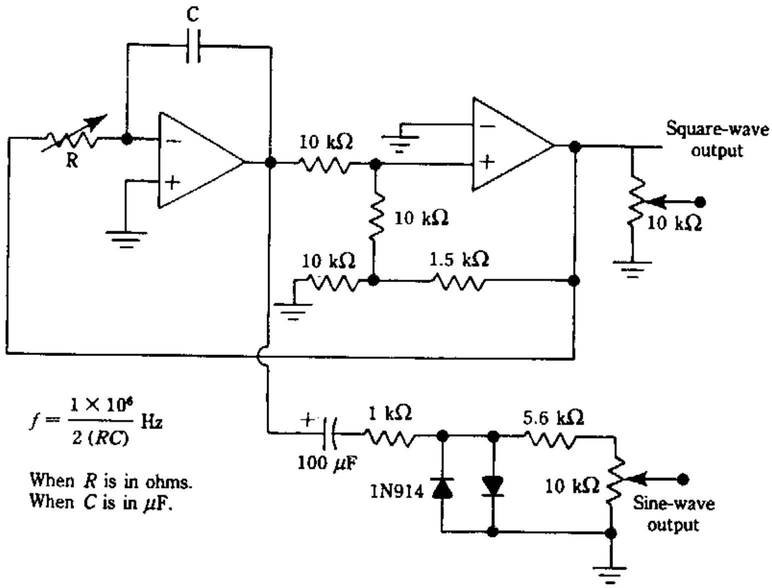
ELECTRONIC DESIGN *Fig. 102-9*

The sine-wave generator starts with an 8-MHz signal and divides it by eight to obtain 1 MHz at C1 (IC1's 2-MHz and 500-kHz outputs can serve as alternate drive signals). Q1 level-shifts the 1-MHz pulses so that they can drive the bipolar circuitry necessary for producing a bipolar output. Synchronous counter IC2 divides 1 MHz by 256 to give the desired output frequency (3906 Hz), and IC3 filters the harmonic frequencies.

The filter's clock is taken from the first divide-by-2 tap of IC2 to assure a 50% duty cycle. IC2 further divides this signal by 128 to ensure that the filter's input signal (1 MHz/256) falls within the flat portion of the filter response.

The output of the switched-capacitor filter resembles a sampled sine wave. It can be smoothed by building a first- or second-order low-pass filter around the otherwise uncommitted output op amp.

SIGNAL SOURCE FOR AUDIO AMPLIFIER/INVERTER

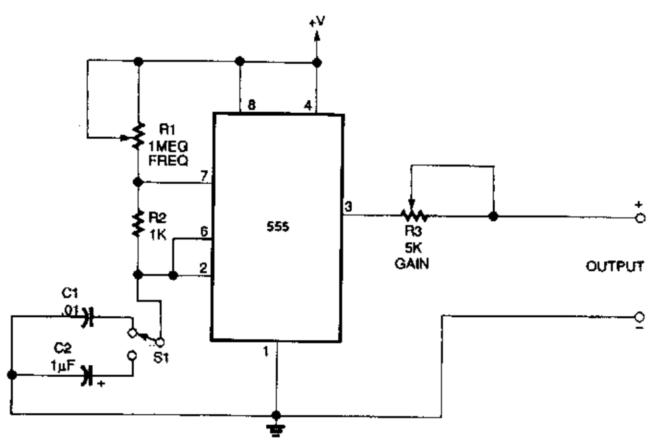


McGRAW-HILL

Fig. 102-10

Two op amps (741, etc..) are used in this oscillator circuit. A square wave is available and a sine wave, obtained by shaping the triangle waveform, is also provided.

SIMPLE TEST SIGNAL GENERATOR



ELECTRONICS NOW

Fig. 102-11

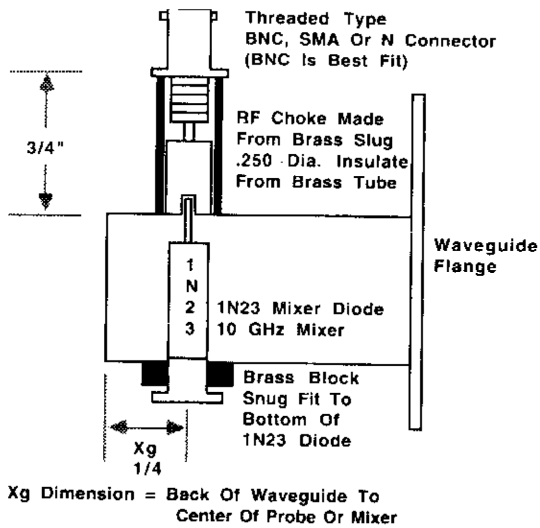
An NE555 generates signals for test purposes. Frequency range is from 20 Hz to 10 kHz, depending on the setting of S1. $+V$ is 5 V.

103

Waveguide Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

10-GHz WAVEGUIDE DETECTOR FOR AMATEUR RADIO USE

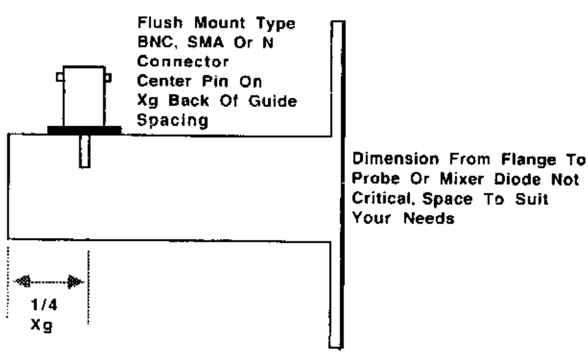


73 AMATEUR RADIO TODAY

Fig. 103-1

This shows the construction of a waveguide detector for use at the 10-GHz amateur radio frequencies.

10-GHz WAVEGUIDE TRANSITION FOR AMATEUR RADIO USE



73 AMATEUR RADIO TODAY

Fig. 103-2

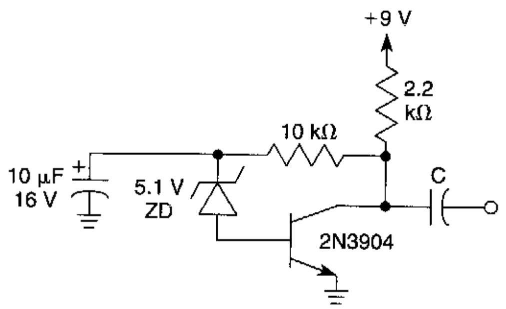
A transistor adapts the waveguide to coaxial cable or other types of transmission lines.

104

White-Noise Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 707. The figure number in the box of each circuit correlates to the entry in the Sources section.

ZENER-DIODE WHITE-NOISE GENERATOR

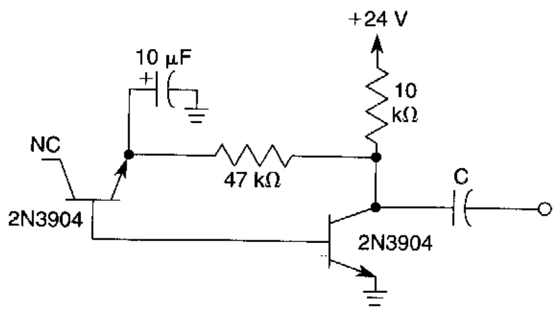


WILLIAM SHEETS

Fig. 104-1

This circuit uses a Zener diode as a noise source. C is chosen to pass the lowest-desired frequency components of the noise.

WHITE-NOISE GENERATOR



WILLIAM SHEETS

Fig. 104-2

Here, a 2N3904 E-B junction is used as a noise generator, reversed bias. C is chosen to pass the lowest-desired frequency components of the noise.

Sources

Chapter 1

Fig. 1-1. QST, 2/95, p. 58.

Fig. 1-2. Analog Devices, Analog Dialogue, Vol. 26, No. 2, p. 5.

Fig. 1-3. Reprinted with permission from Electronic Design, 2/95, p. 108. Copyright 1995, Penton Publishing, Inc.

Fig. 1-4. 1994 Experimenters Handbook, p. 113.

Chapter 2

Fig. 2-1. Reprinted with permission from Electronic Design, 10/94, p. 90. Copyright 1994, Penton Publishing, Inc.

Fig. 2-2. Reprinted with permission from Popular Electronics, 1/94, p. 37. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 2-3. Reprinted with permission from Popular Electronics, Fact Card No. 270. (C) Copyright Gernsback Publications, Inc.

Chapter 3

Fig. 3-1. Reprinted with permission from Popular Electronics, 9/94, p. 26. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 3-2. Reprinted with permission from Popular Electronics, 10/94, p. 48. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 3-3. Reprinted with permission from Popular Electronics, 10/94, p. 35. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 3-4. Reprinted with permission from Popular Electronics, 10/94, p. 49. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 3-5. Reprinted with permission from Popular Electronics, 5/95, pp. 69–70. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 3-6. Reprinted with permission from Popular Electronics, 9/94, p. 24. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 3-7. Reprinted with permission from Popular Electronics, 2/94, p. 37. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 3-8. Reprinted with permission from Popular Electronics, 5/95, p. 70. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 3-9. Reprinted with permission from Popular Electronics, 9/94, p. 25 (C)

Copyright Gernsback Publications, Inc., 1994.

Fig. 3-10. Reprinted with permission from Popular Electronics, 4/95, p. 70. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 3-11. Reprinted with permission from Popular Electronics, 5/95, p. 71. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 4

Fig. 4-1. QST, 12/94, p. 36.

Fig. 4-2. QST, 12/94, p. 35.

Fig. 4-3. QST, 12/94, p. 34.

Fig. 4-4. QST, 5/89, pp. 25–27.

Fig. 4-5. 73 Amateur Radio Today, 5/94, pp. 58–59.

Fig. 4-6. 73 Amateur Radio Today, 8/89, p. 60.

Fig. 4-7. QST, 4/95, p. 56.

Fig. 4-8. QST, 7/94, p. 24.

Fig. 4-9. QST, 3/95, p. 28.

Fig. 4-10. QST, 10/94, p. 65.

Fig. 4-11. 73 Amateur Radio Today, 1/95, p. 32.

Fig. 4-12. 73 Amateur Radio Today, 6/83, p. 99.

Fig. 4-13. 73 Amateur Radio Today, 10/94, p. 14.

Fig. 4-14. 73 Amateur Radio Today, 5/94, p. 10.

Fig. 4-15. William Sheets.

Fig. 4-16. QST, 4/95, p. 61.

Fig. 4-17. QST, 10/94, p. 42.

Fig. 4-18. 73 Amateur Radio Today, 6/94, p. 48.

Fig. 4-19. 73 Amateur Radio Today, 6/94, pp. 32–34.

Chapter 5

Fig. 5-1. Reprinted with permission from Electronics Now, 7/94, p. 34. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 5-2. Reprinted with permission from Electronics Now, 7/94, p. 39. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 5-3. Reprinted with permission from Electronics Now, 7/94, p. 36. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 5-4. William Sheets.

Fig. 5-5. 73 Amateur Radio Today, 9/94, p. 62.

Fig. 5-6. QST, 11/94, p. 23.

Fig. 5-7. QST, 12/94, p. 28.

Fig. 5-8. Rudolf F. Graf and William Sheets.

Fig. 5-9. Reprinted with permission from Radio-Electronics, September 1992, p. 79. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 6

Fig. 6-1. Reprinted with permission from Electronic Design, 11/93, p. 102. Copyright 1993, Penton Publishing, Inc.

Fig. 6-2. Reprinted with permission from Radio-Electronics, Experimenters Handbook, p. 4. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 6-3. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-17.

Fig. 6-4. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-17.

Fig. 6-5. Reprinted with permission from Electronics Now, 2/94, p. 75. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 6-6. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 492.

Fig. 6-7. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, pp. 104–105.

Fig. 6-8. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 107.

Fig. 6-9. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-30.

Fig. 6-10. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-18.

Fig. 6-11. Linear Technology, 2/95.

Fig. 6-12. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-18.

Fig. 6-13. Linear Technology, 2/95.

Fig. 6-14. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 594.

Fig. 6-15. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 127–128.

Fig. 6-16. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 501.

Fig. 6-17. Reprinted with permission from Electronics Now, 6/94, p. 12. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 6-18. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-19.

Chapter 7

Fig. 7-1. Analog Devices, AD8001 Data Sheet.

Fig. 7-2. Reprinted with permission from Popular Electronics, 4/94, p. 47. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 8

Fig. 8-1. 73 Amateur Radio Today, 12/93, p. 32.

Fig. 8-2. Reprinted with permission from Electronic Design, 2/95, p. 107. Copyright 1995, Penton Publishing, Inc.

Fig. 8-3. Reprinted with permission from Electronics Now, 11/93, p. 31. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 8-4. Reprinted with permission from Electronics Now, 5/94, p. 49. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 8-5. QST, 5/95, p. 35.

Fig. 8-6. QST, 5/95, p. 35.

Fig. 8-7. William Sheets.

Chapter 9

Fig. 9-1. QST, 10/94, p. 47.

Fig. 9-2. Reprinted with permission from Popular Electronics, 12/94, p. 88. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 10

Fig. 10-1. Reprinted with permission from Electronics Now, 6/94, p. 34. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 10-2. Reprinted with permission from Electronics Now, 5/93, p. 65. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 10-3. William Sheets.

Fig. 10-4. William Sheets.

Fig. 10-5. Reprinted with permission from Popular Electronics, 12/94, p. 31. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 10-6. William Sheets.

Fig. 10-7. William Sheets.

Fig. 10-8. Reprinted with permission from Electronics Now, 8/94, p. 12. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 10-9. Analog Devices, The Best of Analog Dialogue, 1967–1992, p. 180.

Fig. 10-10. Analog Devices, Analog Dialogue, Vol. 27, No. 2 (1993), p. 22.

Fig. 10-11. Reprinted with permission from Popular Electronics, Fact Card No. 229. (C) Copyright Gernsback Publications, Inc.

Fig. 10-12. William Sheets.

Fig. 10-13. William Sheets.

Fig. 10-14. William Sheets.

Fig. 10-15. William Sheets.

Fig. 10-16. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 180.

Fig. 10-17. William Sheets.

Fig. 10-18. William Sheets.

Fig. 10-19. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 329.

Chapter 11

Fig. 11-1. William Sheets.

Fig. 11-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 1058.

Fig. 11-3. William Sheets.

Fig. 11-4. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 1060.

Fig. 11-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 253.

Fig. 11-6. William Sheets.

Fig. 11-7. Reprinted with permission from Popular Electronics, 1/95, p. 54. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 11-8. Reprinted with permission from

Popular Electronics, 1/95, p. 54. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 11-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 1-14.

Fig. 11-10. William Sheets.

Fig. 11-11. William Sheets.

Fig. 11-12. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 199.

Fig. 11-13. William Sheets.

Fig. 11-14. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 198.

Fig. 11-15. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Edge, #8, p. 14.

Fig. 11-16. William Sheets.

Fig. 11-17. QST, 12/94, p. 44.

Fig. 11-18. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 1059.

Fig. 11-19. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 1059.

Fig. 11-20. Reprinted with permission from Popular Electronics, 5/94, p. 80. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 12

Fig. 12-1. Reprinted with permission from Popular Electronics, 4/95, p. 47. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-2. Reprinted with permission from Popular Electronics, 3/95, p. 67. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-3. William Sheets.

Fig. 12-4. William Sheets.

Fig. 12-5. Reprinted with permission from Popular Electronics, 4/95, p. 70. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-6. Electronics Now, 4/94, p. 25.

Fig. 12-7. Reprinted with permission from Popular Electronics, 5/95, p. 69. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-8. Reprinted with permission from Radio-Electronics, June 1984, p. 39. (C) Copyright Gernsback Publications, Inc., 1984.

Fig. 12-9. Reprinted with permission from Radio-Electronics, 1994 Electronics Experimenters Handbook, p. 37. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 12-10. William Sheets.

Fig. 12-11. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 104.

Fig. 12-12. Reprinted with permission from Popular Electronics, 6/95, p. 32. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-13. Reprinted with permission from Popular Electronics, 4/95, p. 68. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-14. Reprinted with permission from Popular Electronics, 4/95, p. 68. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 12-15. Electronics Now, 4/95, p. 18.

Fig. 12-16. Reprinted with permission from Popular Electronics, 6/95, p. 30. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 13

Fig. 13-1. Reprinted with permission from Electronics Now, 5/95, p. 65. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 13-2. Reprinted with permission from Electronics Now, 10/94, pp. 65–66. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 13-3. William Sheets.

Fig. 13-4. Reprinted with permission from

Electronic Design, 6/94, pp. 42–43. Copyright 1994, Penton Publishing, Inc.

Fig. 13-5. Reprinted with permission from Popular Electronics, 6/95, p. 32. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 13-6. Reprinted with permission from Electronic Design, 6/94, p. 42. Copyright 1994, Penton Publishing, Inc.

Fig. 13-7. Linear Technology, Design Note #98.

Fig. 13-8. 73 Amateur Radio Today, 6/93, p. 41.

Fig. 13-9. 73 Amateur Radio Today, 6/93, p. 35.

Fig. 13-10. Reprinted with permission from Electronic Design, 6/94, p. 42. Copyright 1994, Penton Publishing, Inc.

Fig. 13-11. Reprinted with permission from Popular Electronics, 6/93, p. 76. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 13-12. 73 Amateur Radio Today, 5/93, p. 69.

Chapter 14

Fig. 14-1. Reprinted with permission from Electronic Design, 1/95, pp. 81–82. Copyright 1995, Penton Publishing, Inc.

Fig. 14-2. Spring 1994 Electronics Hobbyist Handbook.

Fig. 14-3. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 499.

Fig. 14-4. Reprinted with permission from Popular Electronics, 3/92, p. 74. (C) Copyright Gernsback Publications, Inc., 1992.

Fig. 14-5. Maxim, Vol. III, New Releases Data Book, 1994, p. 529.

Fig. 14-6. Reprinted with permission from Radio-Electronics, June 1984, p. 90.

Fig. 14-7. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 498.

Fig. 14-8. QST, 10/94, p. 75.

Fig. 14-9. Laser Cookbook, McGraw-Hill, p. 183.

Chapter 15

Fig. 15-1. Reprinted with permission of Burr-Brown Corporation, Burr-Brown Data Sheet INA118, (C) 1989–1995 Burr-Brown Corporation.

Fig. 15-2. Reprinted with permission from Popular Electronics, 3/94, p. 83. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 15-3. Reprinted with permission from Popular Electronics, Fact Card No. 248. (C) Copyright Gernsback Publications, Inc.

Chapter 16

Fig. 16-1. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 180.

Fig. 16-2. William Sheets.

Fig. 16-3. Analog Devices, Analog Dialogue, Vol. 26, No. 2, 1992, p. 17.

Fig. 16-4. Analog Devices, Analog Dialogue, Vol. 26, No. 1, 1992, p. 12.

Fig. 16-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 542.

Fig. 16-6. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 117.

Fig. 16-7. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 543.

Fig. 16-8. Reprinted with permission from Electronic Design, 8/94, p. 104. Copyright 1994, Penton Publishing, Inc.

Fig. 16-9. Linear Technology, Design Note #89.

Chapter 17

Fig. 17-1. Reprinted with permission from Electronic Design, 11/94, p. 115. Copyright 1994, Penton Publishing, Inc.

Fig. 17-2. Reprinted with permission from Electronics Now, 3/95, p. 8. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 18

Fig. 18-1. Reprinted with permission from Electronic Design, 4/89, p. 108. Copyright 1989, Penton Publishing, Inc.

Fig. 18-2. Reprinted with permission from Electronic Design, 10/94, pp. 107–108. Copyright 1994, Penton Publishing, Inc.
Fig. 18-3. 1994 Electronic Experimenters Handbook, p. 99.
Fig. 18-4. Reprinted with permission from Electronic Design, 12/94, p. 115. Copyright 1994, Penton Publishing, Inc.
Fig. 18-5. Maxim, Vol. III, New Releases Data Book, 1994, p. 5-17.
Fig. 18-6. Reprinted with permission from Electronic Design, 10/94, p. 101. Copyright 1994, Penton Publishing, Inc.
Fig. 18-7. Maxim, Vol. III, New Releases Data Book, 1994, p. 4-107.
Fig. 18-8. Reprinted with permission from Electronic Design, 1/84, p. 440. Copyright 1984, Penton Publishing, Inc.
Fig. 18-9. Reprinted with permission from Electronics Now, 5/95, p. 44. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 19

Fig. 19-1. Reprinted with permission from Electronic Design, 3/95, p. 111. Copyright 1995, Penton Publishing, Inc.
Fig. 19-2. Reprinted with permission from Electronics Now, 8/94, p. 8. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 19-3. Reprinted with permission from Electronics Now, 4/95, p. 83. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 19-4. Sound Light and Music, Delton T. Horn, McGraw-Hill, p. 168.

Chapter 20

Fig. 20-1. Radio Craft, 1993, p. 64.
Fig. 20-2. William Sheets.
Fig. 20-3. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 97.
Fig. 20-4. Analog Devices, Analog Dialogue, Vol. 27, No. 2 (1993), p. 21.
Fig. 20-5. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 103.
Fig. 20-6. William Sheets.
Fig. 20-7. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 170–173.
Fig. 20-8. Maxim, Vol. III, New Releases Data Book, 1994, p. 4-79.
Fig. 20-9. Reprinted with permission from Electronic Design, 11/93, p. 89. Copyright 1995, Penton Publishing, Inc.
Fig. 20-10. Reprinted with permission of Burr-Brown Corporation, Burr-Brown Data Sheet INA118, (C) 1989–1995 Burr-Brown Corporation.
Fig. 20-11. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 76. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 20-12. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 76. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 20-13. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, pp. 65–66.
Fig. 20-14. Reprinted with permission from Electronic Design, 3/95, p. 96. Copyright 1995, Penton Publishing, Inc.
Fig. 20-15. Reprinted with permission from Electronic Design, 3/95, pp. 94–96. Copyright 1995, Penton Publishing, Inc.
Fig. 20-16. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.

Chapter 21

Fig. 21-1. Reprinted with permission from Popular Electronics, 2/94, p. 80. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 21-2. Reprinted with permission from Popular Electronics, 2/94, p. 81. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 21-3. Reprinted with permission from Popular Electronics, 2/94, p. 81. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 21-4. Reprinted with permission from Popular Electronics, 2/94, p. 80. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 21-5. William Sheets.
Fig. 21-6. Reprinted with permission from Popular Electronics, 2/94, p. 90. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 21-7. William Sheets.

Fig. 21-8. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 75. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 21-9. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 76. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 21-10. Radio Craft, 1993, p. 63.

Fig. 21-11. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 76. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 22

Fig. 22-1. Reprinted with permission from Popular Electronics, 5/95, pp. 30–31. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 22-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 245.

Fig. 22-3. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 245.

Fig. 22-4. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 245.

Chapter 23

Fig. 23-1. 73 Amateur Radio Today, 3/95, p. 62.

Fig. 23-2. William Sheets.

Fig. 23-3. Linear Technology, Design Note #86.

Fig. 23-4. Maxim, Vol. III, New Releases Data Book, 1994, p. 4-92.

Fig. 23-5. Reprinted with permission from Popular Electronics, 1/94, p. 73. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 24

Fig. 24-1. Reprinted with permission from Electronics Now, 11/93, p. 53. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 24-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 300.

Fig. 24-3. William Sheets.

Fig. 24-4. Reprinted with permission from Electronics Now, 3/95, p. 86. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 24-5. Reprinted with permission from Electronics Now, 3/95, p. 86. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 25

Fig. 25-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook 1991, p. 267.

Fig. 25-2. Reprinted with permission from Electronics Now, 8/93, p. 12. (C) Copyright Gernsback Publications, Inc., 1993.

Chapter 26

Fig. 26-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 541.

Fig. 26-2. Reprinted with permission from Electronic Design, 12/94, p. 129. Copyright 1994, Penton Publishing, Inc.

Fig. 26-3. 1994 Analog Application Issue, Electronic Design, June 27, 1994.

Fig. 26-4. Reprinted with permission from Popular Electronics, 10/94, p. 82. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 26-5. Reprinted with permission from Popular Electronics, 12/94, p. 30. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 26-6. Reprinted with permission from Electronics Now, 3/94, p. 68. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 26-7. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 271.

Fig. 26-8. Reprinted with permission from Popular Electronics, Fact Card No. 249. (C) Copyright Gernsback Publications, Inc.

Fig. 26-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 271.

Fig. 26-10. Reprinted with permission from Electronic Design, 9/94, p. 136. Copyright 1994, Penton Publishing, Inc.

Fig. 26-11. Reprinted with permission from Electronic Design, 1/95, p. 81. Copyright 1995, Penton Publishing, Inc.

Fig. 26-12. Reprinted with permission from Electronics Now, 11/93, p. 8. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 26-13. Reprinted with permission from Electronics Now, 3/94, p. 67. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 26-14. Reprinted with permission from Electronic Design, 5/90, p. 79. Copyright 1990, Penton Publishing, Inc.

Fig. 26-15. Reprinted with permission from Popular Electronics, Fact Card No. 270. (C) Copyright Gernsback Publications, Inc.

Fig. 26-16. Reprinted with permission from Popular Electronics, Fact Card No. 270. (C) Copyright Gernsback Publications, Inc.

Fig. 26-17. Reprinted with permission from Popular Electronics, Fact Card No. 269. (C) Copyright Gernsback Publications, Inc.

Fig. 26-18. Reprinted with permission from Popular Electronics, Fact Card No. 269. (C) Copyright Gernsback Publications, Inc.

Fig. 26-19. Linear Technology, Design Note 88.

Fig. 26-20. Reprinted with permission from Popular Electronics, 10/94, p. 82. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 27

Fig. 27-1. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 37.

Fig. 27-2. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-45.

Chapter 28

Fig. 28-1. Reprinted with permission from Electronics Now, 3/95, p. 64. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 28-2. Reprinted with permission from Popular Electronics, 1/94, p. 24. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 28-3. Reprinted with permission from Popular Electronics, 1/94, p. 25. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 28-4. Linear Technology, Advertisement.

Fig. 28-5. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-127.

Fig. 28-6. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 132–135.

Fig. 28-7. Reprinted with permission from Popular Electronics, 2/94, p. 26. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 28-8. Reprinted with permission from Electronic Design, 12/93, p. 74. Copyright 1993, Penton Publishing, Inc.

Fig. 28-9. Reprinted with permission from Electronics Now, 3/95, p. 61. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 29

Fig. 29-1. Linear Technology, 2/95.

Fig. 29-2. Analog Devices, Analog Dialogue, Vol. 26, No. 2, 1992, p. 18.

Fig. 29-3. Reprinted with permission from Popular Electronics, 10/94, p. 24. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 29-4. Analog Devices, Analog Dialogue, Vol. 26, No. 2, 1992, p. 12.

Fig. 29-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 542.

Fig. 29-6. Reprinted with permission from Popular Electronics, 1/94, p. 24. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 29-7. 73 Amateur Radio Today, 3/95, p. 62.

Fig. 29-8. Reprinted with permission from Popular Electronics, 11/94, p. 31. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 29-9. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), p. 17.

Fig. 29-10. Reprinted with permission from Electronic Design, 7/94, p. 62. Copyright 1994, Penton Publishing, Inc.

Fig. 29-11. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 541.

Fig. 29-12. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), p. 16.

Fig. 29-13. Reprinted with permission from Electronic Design, 9/94, p. 135. Copyright 1994, Penton Publishing, Inc.

Chapter 30

Fig. 30-1. Reprinted with permission from Popular Electronics, 4/95, pp. 29–30. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 30-2. Reprinted with permission from Popular Electronics, 4/95, p. 31. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 30-3. Reprinted with permission from Popular Electronics, 4/95, p. 30. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 31

Fig. 31-1. Laser Cookbook, McGraw-Hill, p. 231.

Fig. 31-2. Laser Cookbook, McGraw-Hill, p. 231.

Chapter 32

Fig. 32-1. Reprinted with permission from Electronic Design, 2/95, p. 115. Copyright 1995, Penton Publishing, Inc.

Fig. 32-2. Amplifiers, Waveform Generators, & Other Low-Cost IC Projects, McGraw-Hill, p. 186.

Fig. 32-3. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 193.

Fig. 32-4. Radio-Electronics Experimenters Handbook, p. 66.

Fig. 32-5. Reprinted with permission from Electronic Design, 12/94, p. 134. Copyright 1994, Penton Publishing, Inc.

Fig. 32-6. Reprinted with permission from Electronic Design, 9/94, p. 79. Copyright 1994, Penton Publishing, Inc.

Fig. 32-7. William Sheets.

Fig. 32-8. Reprinted with permission from Popular Electronics, 12/94, p. 42. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 32-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1202.

Fig. 32-10. Radio Receiver Projects You Can Build, McGraw-Hill, p. 165.

Fig. 32-11. Radio Receiver Projects You Can Build, McGraw-Hill, p. 291.

Fig. 32-12. Reprinted with permission from Electronic Design, 8/94, p. 102. Copyright 1994, Penton Publishing, Inc.

Fig. 32-13. Amplifiers, Waveform, Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 178.

Fig. 32-14. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 229.

Fig. 32-15. Linear Technology, 2/95.

Fig. 32-16. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1187.

Fig. 32-17. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1187.

Fig. 32-18. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 222.

Fig. 32-19. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 79.

Fig. 32-20. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1032.

Fig. 32-21. Linear Technology, Design Note #84.

Fig. 32-22. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1202.

Fig. 32-23. Amplifiers, Waveform Generators, & Other Low-Cost IC Projects, McGraw-Hill, p. 186.

Fig. 32-24. Amplifiers, Waveform Generators, & Other Low-Cost IC Projects, McGraw-Hill, p. 185.

Fig. 32-25. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 201.

Fig. 32-26. Linear Technology, Design Note #89.

Fig. 32-27. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1015.

Fig. 32-28. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 228.

Chapter 33

Fig. 33-1. Reprinted with permission from Popular Electronics, 5/95, pp. 57–59. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 33-2. Reprinted with permission from Popular Electronics, 6/95, p. 77. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 33-3. Reprinted with permission from Popular Electronics, 6/95, p. 78. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 33-4. Reprinted with permission from Popular Electronics, 1/94, p. 73. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 33-5. Sound Light and Music, Delton T. Horn, McGraw-Hill, p. 34.

Fig. 33-6. Reprinted with permission from Popular Electronics, 5/93, p. 70. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 33-7. Reprinted with permission from Popular Electronics, 5/93, p. 70. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 33-8. Reprinted with permission from Popular Electronics, 5/93, p. 72. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 33-9. Reprinted with permission from Popular Electronics, 2/94, p. 25. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 34

Fig. 34-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 241.

Fig. 34-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 241.

Chapter 35

Fig. 35-1. Reprinted with permission from Electronic Design, 2/94, p. 115. Copyright 1994, Penton Publishing, Inc.

Fig. 35-2. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.

Chapter 36

Fig. 36-1. Reprinted with permission from Electronics Now, 4/94, pp. 41–45. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 36-2. Reprinted with permission from Electronics Now, 5/95, pp. 53–55. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 36-3. Reprinted with permission from Popular Electronics, Fact Card No. 249. (C) Copyright Gernsback Publications, Inc.

Fig. 36-4. Reprinted with permission from Electronic Design, 10/94, p. 110. Copyright 1994, Penton Publishing, Inc.

Fig. 36-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 368.

Chapter 37

Fig. 37-1. Reprinted with permission from Popular Electronics, 9/94, p. 81. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-2. Reprinted with permission from Popular Electronics, 9/94, p. 83. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-3. Reprinted with permission from Popular Electronics, 4/94, p. 79. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-4. Reprinted with permission from Popular Electronics, 12/94, pp. 44–47. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-5. Reprinted with permission from Popular Electronics, 4/94, p. 80. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-6. Reprinted with permission from Popular Electronics, 9/94, p. 81. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-7. Reprinted with permission from Popular Electronics, 9/94, pp. 83 and 92. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-8. Reprinted with permission from Popular Electronics, 5/93, p. 62. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 37-9. Reprinted with permission from Popular Electronics, 4/94, p. 78. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-10. Reprinted with permission from Popular Electronics, 4/94, p. 78. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 37-11. Reprinted with permission from Popular Electronics Hobbyist Handbook, 1991, p. 4. (C) Copyright Gernsback Publications, Inc., 1991.

Chapter 38

Fig. 38-1. Reprinted with permission from 1987 Radio-Electronics Experimenters Handbook, p. 63. (C) Copyright Gernsback Publications, Inc., 1987.

Fig. 38-2. Reprinted with permission from Popular Electronics, 1/94, p. 62. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 39

Fig. 39-1. Reprinted with permission from Popular Electronics, 6/93, p. 74. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 39-2. William Sheets.

Fig. 39-3. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), p. 21.

Chapter 40

Fig. 40-1. Reprinted with permission from Popular Electronics, 4/90, p. 102. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 40-2. Reprinted with permission from Popular Electronics, 6/95, p. 76. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 40-3. Reprinted with permission from Popular Electronics, 12/93, p. 31. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 40-4. Reprinted with permission from Popular Electronics, 5/94, p. 69. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 40-5. Reprinted with permission from Popular Electronics, 4/90, p. 90. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 40-6. Reprinted with permission from Popular Electronics, 4/94, p. 34. (C)

Copyright Gernsback Publications, Inc., 1994.

Fig. 40-7. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-100.

Fig. 40-8. Reprinted with permission from Electronics Now, 2/94, p. 38. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 40-9. Reprinted with permission from Popular Electronics, 6/95, p. 76. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 40-10. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 140–142.

Fig. 40-11. Reprinted with permission from Popular Electronics, 4/90, p. 103. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 40-12. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 144–146.

Fig. 40-13. Reprinted with permission from Popular Electronics, 12/93, p. 32. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 40-14. Laser Cookbook, McGraw-Hill, p. 187.

Fig. 40-15. Reprinted with permission from Popular Electronics, 4/90, p. 91. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 40-16. Sound Light and Music, McGraw-Hill.

Fig. 40-17. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 209.

Chapter 41

Fig. 41-1.

Fig. 41-2. Linear Technology, 2/95.

Fig. 41-3. Reprinted with permission from Popular Electronics, Fact Card No. 248. (C) Copyright Gernsback Publications, Inc.

Fig. 41-4. Analog Devices, Analog Dialogue, Vol. 27, No. 2 (1993), p. 17.

Fig. 41-5. Reprinted with permission from Popular Electronics, Fact Card No. 248. (C) Copyright Gernsback Publications, Inc.

Fig. 41-6. Reprinted with permission from Popular Electronics, Fact Card No. 247. (C) Copyright Gernsback Publications, Inc.

Fig. 41-7. Reprinted with permission of Burr-Brown Corporation, Burr-Brown Data Sheet INA118, (C) 1989-1995 Burr-Brown Corporation.

Fig. 41-8. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-33.

Fig. 41-9. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-69.

Fig. 41-10. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), p. 20.

Fig. 41-11. Linear Technology, 2/95.

Chapter 42

Fig. 42-1. Reprinted with permission from Popular Electronics, Fact Card No. 247. (C) Copyright Gernsback Publications, Inc.

Fig. 42-2. Reprinted with permission from Popular Electronics, Fact Card No. 249. (C) Copyright Gernsback Publications, Inc.

Fig. 42-3. William Sheets.

Chapter 43

Fig. 43-1. Reprinted with permission from Popular Electronics, 4/90, p. 78. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 43-2. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.

Chapter 44

Fig. 44-1. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 173–174.

Fig. 44-2. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 126.

Fig. 44-3. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 128.

Fig. 44-4. Maxim, Vol. III, New Releases Data Book, 1994, p. 4-92.

Fig. 44-5. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 152–153.

Chapter 45

Fig. 45-1. Fantastic Electronics, McGraw-Hill, pp. 36–38.

Fig. 45-2. 73 Amateur Radio Today, 7/82, p. 53.

Fig. 45-3. 73 Amateur Radio Today, 7/82, p. 53.

Fig. 45-4. Spring 1994 Electronic Hobbyists Handbook.

Chapter 46

Fig. 46-1. Laser Cookbook, McGraw-Hill, pp. 165–167.

Fig. 46-2. Laser Cookbook, McGraw-Hill, pp. 157–158.

Fig. 46-3. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 125.

Fig. 46-4. Laser Cookbook, McGraw-Hill, p. 201.

Fig. 46-5. Laser Cookbook, McGraw-Hill, p. 190.

Fig. 46-6. Laser Cookbook, McGraw-Hill, p. 200.

Fig. 46-7. Laser Cookbook, McGraw-Hill, p. 170.

Fig. 46-8. Reprinted with permission from Popular Electronics, 6/93, p. 78. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 46-9. Laser Cookbook, McGraw-Hill, p. 169.

Fig. 46-10. Laser Cookbook, McGraw-Hill, p. 168.

Chapter 47

Fig. 47-1. Reprinted with permission from Electronics Now, 5/94, p. 12. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 47-2. William Sheets.

Fig. 47-3. Reprinted with permission from Popular Electronics, 6/95, p. 54. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 47-4. Laser Cookbook, McGraw-Hill, pp. 140–141.

Fig. 47-5. Linear Technology, 2/95.

Fig. 47-6. Reprinted with permission from Electronic Design, 12/93, p. 75. Copyright 1993, Penton Publishing, Inc.

Fig. 47-7. Reprinted with permission from Popular Electronics, 9/94, p. 24. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 47-8. Reprinted with permission from Electronic Design, 10/94, p. 92. Copyright 1994, Penton Publishing, Inc.

Fig. 47-9. Reprinted with permission from Popular Electronics, 4/90, p. 77. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 47-10. Reprinted with permission from Popular Electronics, 9/94, p. 83. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 47-11. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, pp. 82–83.

Fig. 47-12. Reprinted with permission from Electronics Now, 5/94, p. 45. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 47-13. Reprinted with permission from Popular Electronics, 1995, p. 77. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 47-14. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, pp. 589–590.

Fig. 47-15. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 218.

Fig. 47-16. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 219.

Fig. 47-17. Spring 1994 Electronics Hobbyist Handbook.

Fig. 47-18. Spring 1994 Electronics Hobbyist Handbook.

Fig. 47-19. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 493.

Fig. 47-20. Reprinted with permission from Popular Electronics, 6/95, p. 56. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 47-21. Reprinted with permission from Electronics Now, 9/94, p. 66. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 48

Fig. 48-1. NASA Tech Briefs, Spring 1977.

Fig. 48-2. Reprinted with permission from Electronic Design, 12/93, p. 73. Copyright 1993, Penton Publishing, Inc.

Fig. 48-3. Reprinted with permission from Popular Electronics, Fact Card No. 268. (C) Copyright Gernsback Publications, Inc.

Fig. 48-4. William Sheets.

Fig. 48-5. William Sheets.

Chapter 49

Fig. 49-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 520.

Fig. 49-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, pp. 519–520.

Fig. 49-3. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 519.

Fig. 49-4. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 78.

Fig. 49-5. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 79.

Fig. 49-6. Analog Dialogue, Analog Devices, Vol. 26, No. 1, pp. 14–15.

Fig. 49-7. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 78.

Fig. 49-8. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 78.

Fig. 49-9. Reprinted with permission from Electronic Design, 3/95, p. 116. Copyright 1995, Penton Publishing, Inc.

Fig. 49-10. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 27.

Fig. 49-11. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 77.

Chapter 50

Fig. 50-1. Reprinted with permission from Popular Electronics, 4/95, p. 54. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-2. 1994 Electronic Experimenters Handbook, p. 89.

Fig. 50-3. 73 Amateur Radio Today, 7/93, p. 34.

Fig. 50-4. NASA Tech Briefs, November 1993, p. 56.

Fig. 50-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 588.

Fig. 50-6. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, pp. 519–520.

Fig. 50-7. Reprinted with permission from Electronics Now, 3/95, p. 42. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-8. Reprinted with permission from Radio-Electronics, February 1989, p. 64. (C) Copyright Gernsback Publications, Inc., 1989.

Fig. 50-9. Reprinted with permission from Electronics Now, 5/95, p. 10. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-10. Reprinted with permission from Electronics Now, 5/95, p. 10. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-11. Reprinted with permission from Popular Electronics, Fact Card No. 221. (C) Copyright Gernsback Publications, Inc.

Fig. 50-12. Fantastic Electronics, McGraw-Hill, pp. 52–60.

Fig. 50-13. Analog Devices, Analog Dialogue, Vol. 27, No. 2, p. 20.

Fig. 50-14. Electronics Now, 9/94, pp. 73–74.

Fig. 50-15. Reprinted with permission from Popular Electronics, 1/94, pp. 31–36. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-16. Electronics Design, June 27, 1994, p. 33.

Fig. 50-17. Reprinted with permission from Popular Electronics, 10/94, p. 90. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-18. Reprinted with permission from Popular Electronics, 11/94, p. 62. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-19. Radio-Electronics, Feb. 1989, p. 64.

Fig. 50-20. Reprinted with permission from Popular Electronics, 11/94, p. 91. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-21. Reprinted with permission from Electronic Design, 10/94, pp. 102–104. Copyright 1994, Penton Publishing, Inc.

Fig. 50-22. Reprinted with permission from Electronic Design, 12/94, p. 132. Copyright 1994, Penton Publishing, Inc.

Fig. 50-23. Reprinted with permission from Electronics Now, 3/95, p. 63. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-24. Reprinted with permission from Popular Electronics, 5/95, pp. 47–48. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-25. Reprinted with permission from Popular Electronics, 5/95, pp. 47–48. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-26. Reprinted with permission from Popular Electronics, 11/94, p. 38. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-27. Reprinted with permission from Electronic Design, 10/94, pp. 92–94. Copyright 1994, Penton Publishing, Inc.

Fig. 50-28. RF Design, August 1994, p. 78.

Fig. 50-29. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 206.

Fig. 50-30. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 500.

Fig. 50-31. Reprinted with permission from Electronic Design, 11/94, pp. 116–118. Copyright 1994, Penton Publishing, Inc.

Fig. 50-32. Reprinted with permission from Electronic Design, 10/94, pp. 107. Copyright 1994, Penton Publishing, Inc.

Fig. 50-33. QST, 12/94, p. 27.

Fig. 50-34. Reprinted with permission from Electronic Design, 8/94, pp. 109. Copyright 1994, Penton Publishing, Inc.

Fig. 50-35. Reprinted with permission of Burr-Brown Corporation, Burr-Brown Data Sheet INA118, (C) 1989–1995 Burr-Brown Corporation.

Fig. 50-36. William Sheets.

Fig. 50-37. Linear Technology, Design Note 96.

Fig. 50-38. Reprinted with permission from Popular Electronics, 11/94, p. 31. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-39. Reprinted with permission of XICOR, XICOR Data Sheet, p. 7. (C) Copyright XICOR, Inc.

Fig. 50-40. Reprinted with permission from Popular Electronics, Fact Card No. 221. (C) Copyright Gernsback Publications, Inc.

Fig. 50-41. Reprinted with permission from Popular Electronics, 11/94, p. 30. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-42. William Sheets.

Fig. 50-43. QST, 10/94, P. 75.

Fig. 50-44. Reprinted with permission from Popular Electronics, 5/95, p. 69. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 50-45. Reprinted with permission from Popular Electronics, 11/94, p. 33. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 50-46. Reprinted with permission from Radio-Electronics, February 1989, pp. 63–65. (C) Copyright Gernsback Publications, Inc., 1989.

Fig. 50-47. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 45.

Fig. 50-48. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, pp. 40–41.

Chapter 51

Fig. 51-1. Sound Light and Music, Delton T. Horn, McGraw-Hill, p. 123.

Fig. 51-2. Sound Light and Music, Delton T. Horn, McGraw-Hill, p. 120.

Chapter 52

Fig. 52-1. Reprinted with permission from Electronic Design, 5/94, pp. 79–80. Copyright 1994, Penton Publishing, Inc.

Fig. 52-2. Reprinted with permission from Popular Electronics, 1/94, p. 26. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-3. Analog Devices, Analog Dialogue, Vol. 26, No. 1, 1992, p. 15.

Fig. 52-4. Reprinted with permission from Popular Electronics, 8/89, pp. 33–35. (C) Copyright Gernsback Publications, Inc., 1989.

Fig. 52-5. NASA Tech Briefs, March 1995, pp. 39–40.

Fig. 52-6. Reprinted with permission from Radio-Electronics Experimenters Handbook, pp. 98–99. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-7. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 99. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-8. 1994 Electronics Experimenters Handbook, p. 57.

Fig. 52-9. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 119–123.

Fig. 52-10. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, pp. 121–122.

Fig. 52-11. Reprinted with permission from Radio-Electronics Experimenters Handbook, pp. 36–38. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-12. Reprinted with permission from Popular Electronics, 4/95, p. 70. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 52-13. Reprinted with permission from Popular Electronics, 4/90, p. 37. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 52-14. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 175–177.

Fig. 52-15. Linear Technology, Design Note 88.

Fig. 52-16. Fantastic Electronics, McGraw-Hill, p. 59.

Fig. 52-17. 73 Amateur Radio Today, 6/83, p. 99.

Fig. 52-18. Spring 1994 Electronic Hobbyists Handbook.

Fig. 52-19. Reprinted with permission from Electronic Design, 1/95, p. 78. Copyright 1995, Penton Publishing, Inc.

Fig. 52-20. Sound Light and Music, Delton T. Horn, McGraw-Hill, p. 123.

Fig. 52-21. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 140–145.

Fig. 52-22. Reprinted with permission from Electronic Design, 3/95, pp. 111–112. Copyright 1995, Penton Publishing, Inc.

Fig. 52-23. Reprinted with permission from Electronic Design, 3/95, p. 94. Copyright 1995, Penton Publishing, Inc.

Fig. 52-24. Reprinted with permission from Electronic Design, 4/89, p. 107. Copyright 1989, Penton Publishing, Inc.

Fig. 52-25. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 367.

Fig. 52-26. 73 Amateur Radio Today, 6/83, p. 99.

Fig. 52-27. Reprinted with permission from Popular Electronics, 4/94, p. 52. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-28. Reprinted with permission from Electronics Now, 2/94, p. 16. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-29. Reprinted with permission from Electronic Design, 3/95, p. 91. Copyright 1995, Penton Publishing, Inc.

Fig. 52-30. William Sheets.

Fig. 52-31. William Sheets.

Fig. 52-32. Reprinted with permission from Electronic Design, 7/94, p. 96. Copyright 1994, Penton Publishing, Inc.

Fig. 52-33. Reprinted with permission from Electronics Now, 4/94, p. 11. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 52-34. Reprinted with permission from Popular Electronics, 5/95, p. 30. (C)

Copyright Gernsback Publications, Inc., 1994.

Fig. 52-35. Reprinted with permission from Electronic Design, 12/94, pp. 115–116. Copyright 1994, Penton Publishing, Inc.

Chapter 53

Fig. 53-1. William Sheets.

Fig. 53-2. William Sheets.

Chapter 54

Fig. 54-1. Reprinted with permission from Popular Electronics, 10/94, p. 28. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 54-2. Reprinted with permission from Popular Electronics, 10/94, p. 26. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 55

Fig. 55-1. Reprinted with permission from Electronic Design, 10/94, pp. 94–96. Copyright 1994, Penton Publishing, Inc.

Fig. 55-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1033.

Fig. 55-3. NASA Tech Briefs, January 1995, pp. 28–29.

Fig. 55-4. Linear Technology, 2/95.

Fig. 55-5. Reprinted with permission from Popular Electronics, Fact Card No. 249. (C) Copyright Gernsback Publications, Inc.

Fig. 55-6. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.

Fig. 55-7. William Sheets.

Chapter 56

Fig. 56-1. 73 Amateur Radio Today, 5/94, p. 12.

Fig. 56-2. 73 Amateur Radio Today, 6/94, pp. 46–48.

Fig. 56-3. Reprinted with permission from Popular Electronics, 5/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Chapter 57

Fig. 57-1. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 167–169.

Fig. 57-2. Reprinted with permission from Electronic Design, 10/94, pp. 104–105. Copyright 1994, Penton Publishing, Inc.

Fig. 57-3. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 355.

Fig. 57-4. Electronics Now, 5/95, p. 8.

Fig. 57-5. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 166–167.

Fig. 57-6. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 163–166.

Fig. 57-7. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 48.

Fig. 57-8. Reprinted with permission from Electronics Now, 5/94, p. 10. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 57-9. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 357.

Chapter 58

Fig. 58-1. William Sheets.

Fig. 58-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 240.

Fig. 58-3. Reprinted with permission from Popular Electronics, 1/94, p. 72. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 58-4. Reprinted with permission from Popular Electronics, 1/94, p. 73. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 58-5. Reprinted with permission from Popular Electronics, Fact Card No. 268. (C) Copyright Gernsback Publications, Inc.

Fig. 58-6. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 24.

Fig. 58-7. William Sheets.

Chapter 59

Fig. 59-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 1-38.

Fig. 59-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor NSLAH 1991, p. 998.

Fig. 59-3. QST, 10/92, p. 22.

Fig. 59-4. William Sheets.

Fig. 59-5. William Sheets.

Chapter 60

Fig. 60-1. Reprinted with permission from Radio-Electronics, June 1987, p. 69. (C) Copyright Gernsback Publications, Inc., 1987.

Fig. 60-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Handbook, 1991, p. 578.

Fig. 60-3. Reprinted with permission from Radio-Electronics, June 1987, p. 70. (C) Copyright Gernsback Publications, Inc., 1987.

Fig. 60-4. Reprinted with permission from Radio-Electronics, June 1987, p. 75. (C) Copyright Gernsback Publications, Inc., 1987.

Fig. 60-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, P. 576.

Fig. 60-6. Reprinted with permission from Electronic Design, 12/94, p. 118. Copyright 1994, Penton Publishing, Inc.

Fig. 60-7. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Handbook, 1991, p. 1058.

Fig. 60-8. Reprinted with permission from Radio-Electronics, June 1987, p. 75. (C) Copyright Gernsback Publications, Inc., 1987.

Fig. 60-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 491.

Fig. 60-10. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 578.

Fig. 60-11. Analog Devices, Analog Dialogue, Vol. 27, No. 2 (1993), p. 17.

Chapter 61

Fig. 61-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, p. 1209.

Fig. 61-2. William Sheets.

Fig. 61-3. William Sheets.

Fig. 61-4. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 17.

Fig. 61-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, P. 1210.

Fig. 61-6. William Sheets.

Fig. 61-7. Reprinted with permission from Popular Electronics, 11/94, p. 31. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 61-8. William Sheets.

Fig. 61-9. Reprinted with permission from Popular Electronics, 5/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 61-10. William Sheets.

Fig. 61-11. Reprinted with permission from Popular Electronics, 3/94, p. 83. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 61-12. Reprinted with permission from Popular Electronics, 12/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Chapter 62

Fig. 62-1. Reprinted with permission from Popular Electronics, 12/93, p. 70. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 62-2. Reprinted with permission from Popular Electronics, 12/93, p. 68. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 62-3. Reprinted with permission from

Popular Electronics, 12/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 62-4. Reprinted with permission from Popular Electronics, 12/93, p. 70. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 62-5. Reprinted with permission from Electronic Design, 11/94, pp. 130–132. Copyright 1994, Penton Publishing, Inc.

Fig. 62-6. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 79.

Fig. 62-7. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 116.

Fig. 62-8. Reprinted with permission from Electronics Now, 12/93, p. 16. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 62-9. Reprinted with permission from Electronic Design, 7/94, p. 94. Copyright 1994, Penton Publishing, Inc.

Fig. 62-10. QST, 5/95, p. 50.

Chapter 63

Fig. 63-1. William Sheets.

Fig. 63-2. Reprinted with permission from Popular Electronics, 3/94, p. 84. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 63-3. William Sheets.

Fig. 63-4. William Sheets.

Fig. 63-5. QST, 4/95, pp. 38–39.

Fig. 63-6. William Sheets.

Fig. 63-7. Reprinted with permission from Radio-Electronics Experimenters Handbook, p. 75. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 63-8. Radio Receiver Projects You Can Build, McGraw-Hill, p. 237.

Fig. 63-9. William Sheets.

Fig. 63-10. Reprinted with permission from Popular Electronics, 2/94, p. 90. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 63-11. William Sheets.

Fig. 63-12. Reprinted with permission from Popular Electronics, 11/94, p. 42. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 63-13. Reprinted with permission from Popular Electronics, 2/94, p. 90. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 63-14. Reprinted with permission from Electronic Design, 10/94, p. 96. Copyright 1994, Penton Publishing, Inc.

Fig. 63-15. William Sheets.

Fig. 63-16. 73 Amateur Radio Today, 5/94, p. 66.

Fig. 63-17. Radio Craft, 1993, p. 63.

Fig. 63-18. Radio-Electronics Experimenters Handbook, 1992, p. 76.

Fig. 63-19. William Sheets.

Fig. 63-20. William Sheets.

Fig. 63-21. William Sheets.

Chapter 64

Fig. 64-1. Reprinted with permission from Electronics Now, 5/95, pp. 65–66. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 64-2. Reprinted with permission from Electronic Design, 11/94, pp. 118–120. Copyright 1994, Penton Publishing, Inc.

Fig. 64-3. Reprinted with permission from Electronic Design, 9/94, p. 84. Copyright 1994, Penton Publishing, Inc.

Fig. 64-4. 73 Amateur Radio Today, 7/94, p. 39.

Chapter 65

Fig. 65-1. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 107–108.

Fig. 65-2. Reprinted with permission from Electronics Now, 5/95, p. 91. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 65-3. 1994 Electronics Experimenters Handbook, p. 67.

Chapter 66

Fig. 66-1. Reprinted with permission from Popular Electronics, 3/94, p. 70. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 66-2. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 94.

Chapter 67

Fig. 67-1. Reprinted with permission from Electronics Now, 5/95, pp. 67–68. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 67-2. Reprinted with permission from Popular Electronics, 10/94, p. 84. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 67-3. Reprinted with permission from Electronics Now, 3/94, p. 59. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 67-4. Reprinted with permission from Popular Electronics, 12/94, pp. 57–59. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 68

Fig. 68-1. Reprinted with permission from Popular Electronics, 4/95, p. 60. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 68-2. Reprinted with permission from Popular Electronics, 4/95, p. 60. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 68-3. Reprinted with permission from Electronic Design, 11/94, p. 133. Copyright 1994, Penton Publishing, Inc.

Fig. 68-4. Reprinted with permission from Popular Electronics, 4/90, pp. 45–46. (C) Copyright Gernsback Publications, Inc., 1990.

Fig. 68-5. Maxim, Vol. III, New Releases Data Book, 1994, p. 4-131.

Fig. 68-6. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 133.

Fig. 68-7. Linear Technology, Design Note #87.

Fig. 68-8. 1994 Electronics Experimenters Handbook, p. 39.

Fig. 68-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 449.

Fig. 68-10. Linear Technology, Design Note #74.

Fig. 68-11. Reprinted with permission from Radio-Electronics Experimenters Handbook 1992, p. 74. (C) Copyright Gernsback Publications, Inc., 1992.

Fig. 68-12. William Sheets.

Fig. 68-13. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 30.

Fig. 68-14. Reprinted with permission from Electronic Design, 7/94, p. 34. Copyright 1994, Penton Publishing, Inc.

Fig. 68-15. Reprinted with permission from Electronic Design, 6/94, p. 32. Copyright 1994, Penton Publishing, Inc.

Fig. 68-16. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 161–163.

Fig. 68-17. NASA Tech Briefs, August 1994, p. 39.

Fig. 68-18. Linear Technology, Design Note #87.

Fig. 68-19. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 367.

Fig. 68-20. Reprinted with permission from Electronic Design, 12/94, p. 130. Copyright 1994, Penton Publishing, Inc.

Fig. 68-21. Reprinted with permission from Electronic Design, 10/94, pp. 108–110. Copyright 1994, Penton Publishing, Inc.

Fig. 68-22. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1063.

Fig. 68-23. Reprinted with permission from Popular Electronics, 6/95, p. 78. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 68-24. Linear Technology, Design Note #74.

Fig. 68-25. Reprinted with permission from Electronic Design, 9/94, p. 140. Copyright 1994, Penton Publishing, Inc.

Fig. 68-26. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1063.

Fig. 68-27. Reprinted with permission from

Electronic Design, 7/94, p. 30. Copyright 1994, Penton Publishing, Inc.

Fig. 68-28. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 450.

Fig. 68-29. William Sheets.

Fig. 68-30. Maxim, Vol. III, New Releases Data Book, 1994, p. 4-131.

Fig. 68-31. Laser Cookbook, McGraw-Hill, p. 172.

Fig. 68-32. Analog Devices, Analog Dialogue, Vol. 27, No. 2, p. 19.

Fig. 68-33. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 450.

Fig. 68-34. Laser Cookbook, McGraw-Hill, p. 172.

Fig. 68-35. Reprinted with permission from Popular Electronics, 5/95, p. 94. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 68-36. 73 Amateur Radio Today, 5/93, p. 51.

Fig. 68-37. Reprinted with permission from Electronics Now, 12/93, p. 14. (C) Copyright Gernsback Publications, Inc., 1993.

Chapter 69

Fig. 69-1. Reprinted with permission from Popular Electronics, 6/93, p. 77. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 69-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 351.

Fig. 69-3. Reprinted with permission from Electronics Now, 10/94, pp. 58–61. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 69-4. William Sheets.

Fig. 69-5. Reprinted with permission from Popular Electronics, 6/93, p. 77. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 69-6. Laser Cookbook, McGraw-Hill, p. 163.

Fig. 69-7. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 498.

Fig. 69-8. Fantastic Electronics, McGraw-Hill, p. 177.

Fig. 69-9. Gordon McComb's Gadgeteer's Goldmine, p. 269.

Fig. 69-10. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 497.

Fig. 69-11. Laser Cookbook, McGraw-Hill, pp. 160–161.

Chapter 70

Fig. 70-1. 73 Amateur Radio Today, 5/93, p. 32.

Fig. 70-2. Laser Cookbook, McGraw-Hill, p. 172.

Fig. 70-3. Reprinted with permission from Electronic Design, 5/90, p. 80. Copyright 1990, Penton Publishing, Inc.

Fig. 70-4. Linear Technology, Design Note #99.

Fig. 70-5. Radio Receiver Projects You Can Build, McGraw-Hill 4256, p. 241.

Fig. 70-6. Reprinted with permission from Popular Electronics, 11/94, p. 41. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 70-7. Linear Technology, Design Note #78.

Fig. 70-8. Reprinted with permission from Electronics Now, 2/94, p. 83. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 70-9. Reprinted with permission from Radio-Electronics, June 1987, p. 75. (C) Copyright Gernsback Publications, Inc., 1987.

Chapter 71

Fig. 71-1. Reprinted with permission from Popular Electronics, 4/95, p. 96. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 71-2. Reprinted with permission from Electronic Design, 6/94, p. 62. Copyright 1994, Penton Publishing, Inc.

Fig. 71-3. QST, 10/92, p. 50.
Fig. 71-4. William Sheets.
Fig. 71-5. Reprinted with permission from Popular Electronics, 4/95, p. 96. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 71-6. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 450.
Fig. 71-7. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 495.
Fig. 71-8. William Sheets.

Chapter 72

Fig. 72-1. Reprinted with permission from Electronic Design, 1/95, p. 133. Copyright 1995, Penton Publishing, Inc.
Fig. 72-2. Reprinted with permission from Electronics Now, 3/95, p. 64. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 72-3. 73 Amateur Radio Today, 6/83, p. 99.
Fig. 72-4. Reprinted with permission from Popular Electronics, 2/94, p. 58. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 72-5. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), p. 20.
Fig. 72-6. Reprinted with permission from Radio-Electronics Experimenters Handbook, pp. 77–82. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 73

Fig. 73-1. Reprinted with permission from Popular Electronics, 11/94, pp. 31 and 91. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 73-2. William Sheets.

Chapter 74

Fig. 74-1. Reprinted with permission from Popular Electronics, 6/95, p. 38. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 74-2. Reprinted with permission from Popular Electronics, 6/95, p. 39. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 75

Fig. 75-1. Fantastic Electronics, McGraw-Hill, pp. 67–73.
Fig. 75-2. Reprinted with permission from Electronics Now, 1/94, p. 61. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 75-3. Reprinted with permission from Electronics Now, 1/94, p. 61. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 75-4. Reprinted with permission from Electronics Now, 1/94, p. 59. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 75-5. Reprinted with permission from Popular Electronics, 5/93, p. 40. (C) Copyright Gernsback Publications, Inc., 1993.

Chapter 76

Fig. 76-1. Reprinted with permission from Popular Electronics, 5/95, p. 66. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 76-2. Reprinted with permission from Popular Electronics, 1/95, p. 49. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 76-3. Radio Receiver Projects You Can Build, McGraw-Hill 4256, pp. 122–129.
Fig. 76-4. Radio-Electronics Experimenters Handbook, p. 65.
Fig. 76-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 2-38.
Fig. 76-6. Radio Craft, 1993, p. 50.
Fig. 76-7. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), pp. 15-16.
Fig. 76-8. Reprinted with permission from Popular Electronics, 3/94, p. 35. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-9. Radio Receiver Projects You Can Build, McGraw-Hill, pp. 158–165.
Fig. 76-10. Radio Receiver Projects You Can Build, McGraw-Hill, pp. 138–147.

Fig. 76-11. Radio Receiver Projects You Can Build, McGraw-Hill, pp. 102–109.
Fig. 76-12. Radio Receiver Projects You Can Build, McGraw-Hill, pp. 196–203.
Fig. 76-13. Radio Receiver Projects You Can Build, McGraw-Hill, pp. 32–41.
Fig. 76-14. Analog Devices, Analog Dialogue, Vol. 26, No. 2, 1992, p. 19.
Fig. 76-15. Reprinted with permission from Popular Electronics, 5/94, p. 79. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-16. Reprinted with permission from Popular Electronics, 10/94, p. 25. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-17. Reprinted with permission from Electronics Now, 3/94, p. 68. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-18. Reprinted with permission from Electronics Now, 3/94, p. 67. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-19. Reprinted with permission from Popular Electronics, 10/94, p. 62. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-20. Analog Devices, Analog Dialogue, Vol. 27, No. 1 (1993), p. 15.
Fig. 76-21. Radio Craft, 1993, p. 64.
Fig. 76-22. William Sheets.
Fig. 76-23. Reprinted with permission from Electronics Now, 3/94, p. 69. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-24. Reprinted with permission from Electronics Now, 3/94, p. 69. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-25. Reprinted with permission from Electronics Now, 3/94, pp. 70–72. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 76-26. Reprinted with permission from Popular Electronics, 5/95, pp. 55-56. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 76-27. William Sheets.
Fig. 76-28. Reprinted with permission from Electronics Now, 3/95, p. 50. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 77

Fig. 77-1. Reprinted with permission from Electronic Design, 6/94, p. 29. Copyright 1995, Penton Publishing, Inc.
Fig. 77-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 26.
Fig. 77-3. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 26.

Chapter 78

Fig. 78-1. Linear Technology, Design Note #98.
Fig. 78-2. Reprinted with permission from Electronics Now, 7/94, p. 12. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 78-3. Reprinted with permission from Electronic Design, 2/95, p. 118. Copyright 1995, Penton Publishing, Inc.
Fig. 78-4. Linear Technology, Design Note #100.
Fig. 78-5. Linear Technology, Design Note #98.
Fig. 78-6. NASA Tech Briefs, August 1994, p. 38.
Fig. 78-7. Linear Technology, Advertisement LT/1294.
Fig. 78-8. Reprinted with permission from Electronic Design, 7/94, p. 32. Copyright 1994, Penton Publishing, Inc.
Fig. 78-9. Reprinted with permission from Electronic Design, 1/95, pp. 133–134. Copyright 1995, Penton Publishing, Inc.
Fig. 78-10. Linear Technology, Design Note #98.

Chapter 79

Fig. 79-1. William Sheets.
Fig. 79-2. William Sheets.
Fig. 79-3. William Sheets.
Fig. 79-4. William Sheets.
Fig. 79-5. William Sheets.
Fig. 79-6. William Sheets.
Fig. 79-7. William Sheets.

Chapter 80

Fig. 80-1. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-30.

Fig. 80-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1202.

Chapter 81

Fig. 81-1. Reprinted with permission from Popular Electronics, 11/94, pp. 75–76. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 81-2. Reprinted with permission from Popular Electronics, 11/94, pp. 74–75. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 82

Fig. 82-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 1-37.

Fig. 82-2. Reprinted with permission from Popular Electronics, 9/94, p. 73. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 82-3. 73 Amateur Radio Today, 4/95, pp. 54–58.

Chapter 83

Fig. 83-1. William Sheets.

Fig. 83-2. William Sheets.

Fig. 83-3. Reprinted with permission from Popular Electronics, 5/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 83-4. William Sheets.

Fig. 83-5. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 105–106.

Chapter 84

Fig. 84-1. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 106–110.

Fig. 84-2. Reprinted with permission from Popular Electronics, 1/94, p. 75. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 84-3. 1994 Electronic Experimenters Handbook, p. 53.

Fig. 84-4. William Sheets.

Fig. 84-5. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 98–100.

Fig. 84-6. Sound Light and Music, Delton T. Horn, McGraw-Hill, pp. 114–115.

Fig. 84-7. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 147.

Fig. 84-8. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 127.

Chapter 85

Fig. 85-1. Reprinted with permission from Popular Electronics, Fact Card No. 221. (C) Copyright Gernsback Publications, Inc.

Fig. 85-2. William Sheets.

Chapter 86

Fig. 86-1. Reprinted with permission from Popular Electronics, 11/94, pp. 76 and 91. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 86-2. William Sheets.

Fig. 86-3. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 235.

Fig. 86-4. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 236.

Chapter 87

Fig. 87-1. Reprinted with permission from Electronic Design, 7/94, p. 94. Copyright 1994, Penton Publishing, Inc.

Fig. 87-2. Reprinted with permission from Electronics Now, 2/94, p. 55. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 87-3. Reprinted with permission from Electronics Now, 2/94, p. 57. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 88

Fig. 88-1. Reprinted with permission from Popular Electronics, 6/93, p. 71. (C)

Copyright Gernsback Publications, Inc., 1993.

Fig. 88-2. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 117.

Fig. 88-3. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 117.

Fig. 88-4. Reprinted with permission from Electronic Design, 7/94, pp. 96–97. Copyright 1994, Penton Publishing, Inc.

Fig. 88-5. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 117.

Fig. 88-6. Reprinted with permission from Popular Electronics, 6/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 88-7. Reprinted with permission from Electronics Now, 5/95, p. 63. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 88-8. Reprinted with permission from Electronic Design, 1/94, p. 118. Copyright 1994, Penton Publishing, Inc.

Fig. 88-9. Reprinted with permission from Popular Electronics, 12/94, p. 41. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 88-10. Reprinted with permission from Popular Electronics, 12/94, p. 42. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 88-11. Reprinted with permission from Popular Electronics, 12/94, p. 41. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 88-12. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 72.

Fig. 88-13. Reprinted with permission from Popular Electronics, 12/94, p. 41. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 88-14. Reprinted with permission from Popular Electronics, 12/94, p. 41. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 88-15. Reprinted with permission from Popular Electronics, 12/94, p. 41. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 88-16. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-125.

Fig. 88-17. William Sheets.

Fig. 88-18. Reprinted with permission from Popular Electronics, 6/93, p. 71. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 88-19. William Sheets.

Fig. 88-20. Reprinted with permission from Popular Electronics, 6/93, p. 72. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 88-21. Electronics Now, 8/93, p. 12.

Fig. 88-22. William Sheets.

Fig. 88-23. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 117.

Chapter 89

Fig. 89-1. Reprinted with permission from Electronic Design, 10/88, pp. 126–128. Copyright 1988, Penton Publishing, Inc.

Fig. 89-2. Analog Devices, Analog Dialogue, Vol. 26, No. 1, 1992, p. 12.

Fig. 89-3. Reprinted with permission from Electronic Design, 11/93, p. 99. Copyright 1993, Penton Publishing, Inc.

Fig. 89-4. Analog Devices, Analog Dialogue, Vol. 26, No. 1, 1992, p. 13.

Fig. 89-5. Reprinted with permission from Electronic Design, 10/88, p. 128. Copyright 1988, Penton Publishing, Inc.

Chapter 90

Fig. 90-1. Reprinted with permission from 1987 Radio-Electronics Experimenters Handbook, p. 51. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 90-2. Reprinted with permission from Electronics Now, 2/94, p. 33. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 90-3. Reprinted with permission from Electronics Now, 10/94, p. 53. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 90-4. 1994 Electronics Experimenters Handbook, p. 123.

Fig. 90-5. Reprinted with permission from Electronics Now, 4/95, pp. 39–40. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 90-6. Reprinted with permission from Electronics Now, 8/94, p. 26. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 90-7. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.
Fig. 90-8. Reprinted with permission from Popular Electronics, 12/93, p. 62. (C) Copyright Gernsback Publications, Inc., 1993.
Fig. 90-9. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, pp. 335–336.
Fig. 90-10. Reprinted with permission from Electronics Now, 2/94, p. 16. (C) Copyright Gernsback Publications, Inc., 1994.
Fig. 90-11. Reprinted with permission from Electronics Now, 8/93, pp. 58–63. (C) Copyright Gernsback Publications, Inc., 1993.
Fig. 90-12. Reprinted with permission from Radio-Electronics, August 1993, pp. 58–63. (C) Copyright Gernsback Publications, Inc., 1993.
Fig. 90-13. Reprinted with permission from Electronic Design, 8/94, p. 116. Copyright 1994, Penton Publishing, Inc.
Fig. 90-14. William Sheets.

Chapter 91

Fig. 91-1. 1994 Electronics Experimenters Handbook, p. 63.
Fig. 91-2. Reprinted with permission from Electronic Design, 7/94, p. 93. Copyright 1994, Penton Publishing, Inc.
Fig. 91-3. Reprinted with permission from Electronic Design, 11/93, pp. 90–92. Copyright 1993, Penton Publishing, Inc.
Fig. 91-4. Reprinted with permission from Electronic Design, 1/95, pp. 80–81. Copyright 1995, Penton Publishing, Inc.
Fig. 91-5. Reprinted with permission from Popular Electronics, 6/95, p. 48. (C) Copyright Gernsback Publications, Inc., 1995.
Fig. 91-6. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 524.
Fig. 91-7. Reprinted with permission from Electronic Design, 1/94, pp. 118–119. Copyright 1994, Penton Publishing, Inc.
Fig. 91-8. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 524.
Fig. 91-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 500.
Fig. 91-10. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1076.
Fig. 91-11. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1076.
Fig. 91-12. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1076.
Fig. 91-13. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1076.
Fig. 91-14. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.
Fig. 91-15. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 75.
Fig. 91-16. Reprinted with permission of National Semiconductor Corporation, National Semiconductor, NSLAH 1991, p. 1079.
Fig. 91-17. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 493.
Fig. 91-18. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1077.
Fig. 91-19. Reprinted with permission from Popular Electronics, 5/93, p. 43. (C) Copyright Gernsback Publications, Inc., 1993.
Fig. 91-20. Reprinted with permission of Burr-Brown Corporation, Burr-Brown Data Sheet INA118, (C) 1989–1995 Burr-Brown Corporation.

Fig. 91-21. Analog Devices, Analog Dialogue, Vol. 27, No. 2, p. 22.

Fig. 91-22. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1076.

Fig. 91-23. QST, 5/95, p. 85.

Fig. 91-24. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 523.

Fig. 91-25. Reprinted with permission from Popular Electronics, 1/95, p. 28. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 91-26. QST, 5/95, p. 85.

Fig. 91-27. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1079.

Fig. 91-28. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 1076.

Fig. 91-29. Reprinted with permission from Popular Electronics, 1/95, p. 29. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 91-30. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, pp. 163–164.

Chapter 92

Fig. 92-1. Reprinted with permission from Popular Electronics, 9/94, p. 26. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 92-2. Reprinted with permission from Popular Electronics, 2/94, p. 25. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 92-3. Reprinted with permission from Electronic Design, 3/95, pp. 114–116. Copyright 1995, Penton Publishing, Inc.

Chapter 93

Fig. 93-1. William Sheets.

Fig. 93-2. William Sheets.

Fig. 93-3. William Sheets.

Fig. 93-4. William Sheets.

Chapter 94

Fig. 94-1. William Sheets.

Fig. 94-2. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 92.

Fig. 94-3. NASA Tech Briefs, August 1994, pp. 34–35.

Fig. 94-4. William Sheets.

Fig. 94-5. Reprinted with permission from Popular Electronics, Fact Card No. 270. (C) Copyright Gernsback Publications, Inc.

Chapter 95

Fig. 95-1. Spring 1994 Electronics Hobbyist Handbook.

Fig. 95-2. Reprinted with permission from Electronics Now, 5/95, p. 8. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 95-3. Reprinted with permission from Popular Electronics, Fact Card No. 221. (C) Copyright Gernsback Publications, Inc.

Chapter 96

Fig. 96-1. Reprinted with permission from Electronics Now, 6/93, pp. 41–46. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 96-2. Reprinted with permission from Electronics Now, 12/93, p. 29. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 96-3. 1994 Electronics Experimenters Handbook, p. 105.

Fig. 96-4. Reprinted with permission from Electronics Now, 1994 Experimenters Handbook, p. 77. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 96-5. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, pp. 340–341.

Fig. 96-6. 73 Amateur Radio Today, 5/93, p. 56.

Fig. 96-7. Reprinted with permission from Popular Electronics, 8/92, p. 45. (C) Copyright Gernsback Publications, Inc., 1992.

Chapter 97

Fig. 97-1. Reprinted with permission from Popular Electronics, 5/95, p. 29. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 97-2. Reprinted with permission from Popular Electronics, 1/95, pp. 72–73. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 97-3. Reprinted with permission from Popular Electronics, 1/95, p. 71. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 97-4. Reprinted with permission from Popular Electronics, 1/95, p. 73. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 97-5. Reprinted with permission from Popular Electronics, 12/94, p. 72. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 97-6. Reprinted with permission from Popular Electronics, 12/94, pp. 74–75. (C) Copyright Gernsback Publications, Inc., 1994.

Chapter 98

Fig. 98-1. Reprinted with permission from Electronics Now, 5/95, pp. 50–51. (C) Copyright Gernsback Publications, Inc., 1995.

Fig. 98-2. Analog Devices, AD8001 Data Sheet.

Fig. 98-3. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 3-50.

Fig. 98-4. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 3-33.

Fig. 98-5. Reprinted with permission from Radio-Electronics Experimenters Handbook, pp. 83–88, 1994. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 98-6. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, p. 1021.

Fig. 98-7. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-16.

Fig. 98-8. 73 Amateur Radio Today, 7/93, p. 74.

Fig. 98-9. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 3-32.

Fig. 98-10. 73 Amateur Radio Today, 7/93, p. 74.

Fig. 98-11. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 3-13.

Fig. 98-12. Linear Technology, Design Note #92.

Fig. 98-13. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-16.

Fig. 98-14. Reprinted with permission from Popular Electronics, Fact Card No. 268. (C) Copyright Gernsback Publications, Inc.

Fig. 98-15. Maxim, Vol. III, New Releases Data Book, 1994, p. 8-19.

Fig. 98-16. Linear Technology, 2/95.

Fig. 98-17. Analog Dialogue, Analog Devices, Vol. 26, No. 1, 1992, p. 11.

Chapter 99

Fig. 99-1. Analog Devices, The Best of Analog Dialogue, 1967–1991, p.75.

Fig. 99-2. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Application Specific IC's Databook 1993, p. 1-37.

Chapter 100

Fig. 100-1. Reprinted with permission from Popular Electronics, Fact Card No. 269. (C) Copyright Gernsback Publications, Inc.

Fig. 100-2. Analog Devices, The Best of Analog Dialogue, 1967–1991, p. 79.

Chapter 101

Fig. 101-1. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 500.

Fig. 101-2. Maxim, Vol. III, New Releases Data Book, 1994, p. 3-126.

Fig. 101-3. Reprinted with permission from Electronic Design, 11/94, pp. 127–128. Copyright 1994, Penton Publishing, Inc.

Fig. 101-4. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 207.

Fig. 101-5. QST, 4/95, p. 61.

Fig. 101-6. 73 Amateur Radio Today, 3/95, p 63.

Fig. 101-7. Reprinted with permission from Popular Electronics, 5/94, p. 80. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 101-8. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 499.

Fig. 101-9. Reprinted with permission from Popular Electronics, 5/94, p. 79. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 101-10. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 206.

Chapter 102

Fig. 102-1. Reprinted with permission from Popular Electronics, 11/94, p. 74. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 102-2. Reprinted with permission from Popular Electronics, 5/94, p. 90. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 102-3. Reprinted with permission from Electronic Design, 12/94, pp. 118–119. Copyright 1994, Penton Publishing, Inc.

Fig. 102-4. Reprinted with permission from Electronics Now, 8/93, p. 14. (C) Copyright Gernsback Publications, Inc., 1993.

Fig. 102-5. Reprinted with permission of National Semiconductor Corporation, National Semiconductor Linear Applications Handbook, 1991, p. 208.

Fig. 102-6. Gordon McComb's Gadgeteer's Goldmine, McGraw-Hill, p. 197.

Fig. 102-7. Amplifiers, Waveform Generators & Other Low-Cost IC Projects, McGraw-Hill, p. 21.

Fig. 102-8. Reprinted with permission from Popular Electronics, 11/94, pp. 76 and 91. (C) Copyright Gernsback Publications, Inc., 1994.

Fig. 102-9. Reprinted with permission from Electronic Design, 7/94, p. 96. Copyright 1994, Penton Publishing, Inc.

Fig. 102-10. Power Supplies, Switching Regulators, Inverters, and Converters, McGraw-Hill, p. 129.

Fig. 102-11. Reprinted with permission from Electronics Now, 3/95, p. 63. (C) Copyright Gernsback Publications, Inc., 1995.

Chapter 103

Fig. 103-1. 73 Amateur Radio Today, 12/93, p. 70.

Fig. 103-2. 73 Amateur Radio Today, 12/93, p. 72.

Chapter 104

Fig. 104-1. William Sheets.

Fig. 104-2. William Sheets.

Index

Numbers preceded by an "I", "II", "III", "IV", "V", or "VI" are from *Encyclopedia of Electronic Circuits* Vol. I, II, III, IV, V, or VI respectively.

A

C

D

F

G

J

K

L

P

Q

R

S

ABOUT THE AUTHORS

Rudolf F. Graf has 45 years of engineering, sales, and marketing experience in the electronics field. He has written more than 30 books (about three million copies printed) and well over 100 articles. He is a senior member of the IEEE, a licensed amateur radio operator (KA2CWL), and has a BSEE degree from Polytechnic Institute of Brooklyn and an MBA from NYU. He is self-employed.

William Sheets is a self-employed circuit design engineer. He has more than 25 years of experience in RF, analog, and digital electronics. He has written numerous articles in electronics publications and co-authored five books with Graf. His interests include amateur radio (K2MQJ), photography, and travel. He has designed and built numerous items, including a satellite TV system, many transmitters and receivers, and a computer. He has an MEE degree from NYU, is married, and lives in upstate New York.