MICROWAVE OSCILLATORS

Oscillators are an integral part of receiving and transmitting systems used for communication, radar and other applications. Any oscillator circuit consists of an active device and a resonant circuit which determines the frequency of oscillation. The active device can be either a two-terminal negative resistance device such as a Gunn diode, an impact avalanche transit time (IMPATT) diode, a resonant tunneling diode (RTD), and so on, or a transistor with appropriate feedback to cause instability. In general, an oscillator can operate at a fixed frequency, or its frequency of operation can be tunable. The frequency tuning can be achieved either mechanically or electronically. An oscillator is usually characterized by its frequency of operation, output power, phase noise, long-term frequency stability, and dc to RF efficiency.

MODELING OF MICROWAVE OSCILLATORS

An oscillator consists of linear circuits and active devices and its operation relies on the nonlinear behavior of the active devices. In order to accurately characterize an oscillator, it has to be analyzed using linear and nonlinear circuit analysis techniques. An oscillator can be analyzed using either the

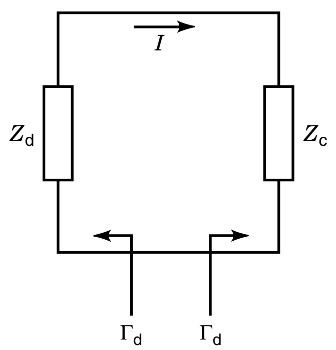

Figure 1. Simplified block diagram of a one-port oscillator.

feedback model or the negative resistance model. These two analysis models are identical; however, the negative resistance model is the most commonly used approach in the design of microwave oscillators. The popularity of the negative resistance model is due to its simplicity and the fact that it can be related to device and circuit reflection coefficients since they can be measured accurately at microwave frequencies. In either model, the oscillator is divided into two parts, the active device and the embedding passive circuit. In a simplified oscillator analysis, the active devide is assumed to operate under small signal conditions. This type of analysis can provide some information about startup condition for oscillation and predict the approximate frequency of oscillation. In order to accurately determine the oscillation frequency, output power, stability, and spectral purity of an oscillator, large signal analysis of the oscillator circuit must be performed. Usually simplified large signal models for active devices can provide valuable insight into the operation of oscillators.

Oscillator Design Using One-Port Negative Resistance Devices

Figure 1 shows the simplified block diagram of a one port oscillator circuit. The small signal impedance of the negative resistance device and the embedding circuit impedance can be denoted as $Z_d = -R_d + jX_d$ and $Z_c = R_c + jX_c$. For a series resonant circuit assuming that current I is flowing through the circuit we obtain

$$I[Z_d + Z_c] = 0 \tag{1}$$

which results in

$$Z_d + Z_c = 0 \tag{2}$$

The startup conditions for oscillation for a series resonant circuit are given by Eqs. (3) and (4):

$$-R_d + R_c < 0 \tag{3}$$

$$X_d + X_c = 0 \tag{4}$$

This means that the magnitude of the device negative resistance must be larger than the overall circuit losses in order for oscillation to build up. The startup of oscillation is initiated either from noise present in the circuit or from transient introduced during power turn-on. As the oscillation amplitude builds up, the device enters into its nonlinear region where the device negative resistance saturates and eventually becomes equal to the overall circuit loses. At this point the oscillation amplitude reaches its steady state.

For a parallel resonant circuit, the steady-state oscillation condition is given in terms of active device and circuit admittances Y_d and Y_c, respectively, as

$$Y_d + Y_c = 0 \tag{5}$$

The startup conditions for a parallel resonant circuit are given by Eqs. (6) and (7):

$$-G_d + G_c < 0 \tag{6}$$

$$B_d + B_c = 0 \tag{7}$$

where $Y_d = -G_d + jB_d$ and $Y_c = G_c + jB_c$.

The steady-state oscillation condition can be written in terms of device and circuit reflection coefficients Γ_d and Γ_c as

$$\Gamma_d \cdot \Gamma_c = 1 \tag{8}$$

Voltage-controlled devices such as a Gunn diode should be connected to a parallel resonant circuit, and the appropriate startup conditions for oscillation should be satisfied. Current-controlled devices such as IMPATT diode are placed in a series resonant circuit for proper operation. In this case, startup conditions for oscillation for a series resonant circuit should be met. The startup oscillation frequency can be different than the steady-state oscillation frequency. This is because the large signal device impedance is a function of the voltage (or current) amplitude across (through) the device. As the oscillation amplitude builds up, the device impedance varies until it reaches its final value. However, for high Q circuits, the steady-state oscillation frequencies are very close to the startup frequency of oscillation. It also should be mentioned that the startup condition for oscillation is a necessary condition but not sufficient condition for having an unstable circuit (1–3). (Nyquist or root locus analysis can always be used to determine the circuit instability.)

Oscillator Design Using Two-Port Devices

The block diagram of a two-port oscillator circuit is shown in Fig. 2. Most microwave transistors are conditionally stable within a limited range of frequencies. By plotting the input and output stability circles on a Smith chart, one can graphically determine the range of impedances for unstable operation. When designing microwave oscillators, the input port of the two-terminal device can be terminated with a purely reac-

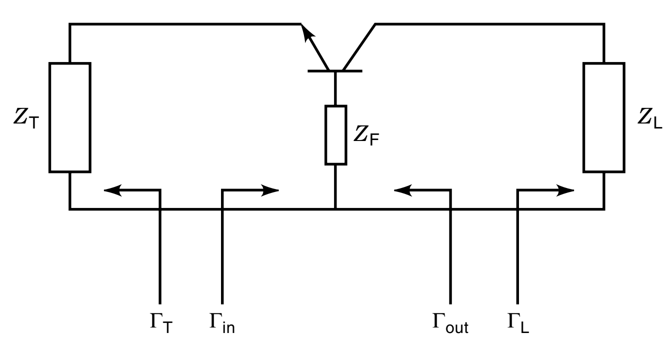

Figure 2. Block diagram of a transistor oscillator.

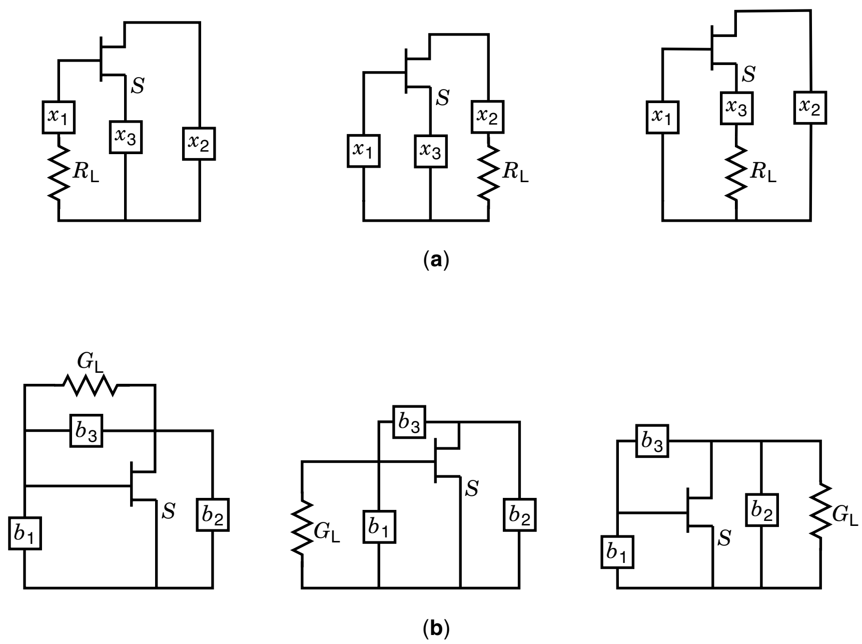

Figure 3. Circuit configurations of (a) series feedback oscillators and (b) shunt feedback oscillators.

tive load that lies in the unstable region of the Smith chart. This is shown in Fig. 2 as Z_T. This will result in the input and output impedances looking into the device to have negative real parts (magnitude of reflection coefficients larger than one). At this point the transistor oscillator can be treated as a one-port negative resistance device. The load impedance Z_L must be chosen to satisfy the startup conditions for oscillation. The steady-state condition for oscillation for a two-port oscillator in terms of the transistor's S-parameters load and source reflection coefficients is given by Eqs. (9) and (10):

$$\frac{1}{\Gamma_T} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \tag{9}$$

$$\frac{1}{\Gamma_L} = S_{22} + \frac{S_{12}S_{21}\Gamma_T}{1 - S_{11}\Gamma_T} \tag{10}$$

If the oscillation conditions are satisfied at one port, they are satisfied at all other ports (4,5).

If the oscillator circuit is stable at the desired frequency of operation, it can be made unstable by using series or shunt feedback (shown as Z_F in Fig. 2). There are a total of six configurations, three for series feedback and three for shunt feedback as shown in Fig. 3 (6).

Large Signal Analysis and Stability

In general the active device's impedance can be expressed as

$$Z_d(A, \omega) = -R_d(A, \omega) + jK_d(A, \omega) \tag{11}$$

where A is the amplitude of current through the active device and ω is the angular frequency. It is assumed that only the voltage across the device contains harmonic components.

Furthermore, we can express the embedding circuit's impedance as

$$\boldsymbol{Z}_c(\omega) = R_c(\omega) + jX_c(\omega) \tag{12}$$

Assuming that the current through the device is nearly sinusoidal—that is, $i(t) = A\cos(\omega t)$—Eq. (13) should be satisfied:

$$A[\boldsymbol{Z}_d(A, \omega) + Z_c(\omega)] = 0 \tag{13}$$

Equation (13) can be rewritten as

$$A[-R_d(A, \omega) + R_c(\omega)] = 0 \tag{14}$$

and

$$A[X_d(A, \omega) + X_c(\omega) = 0 \tag{15}$$

By solving Eqs. (14) and (15), the steady-state amplitude and the frequency of oscillation can be determined. Now assume that the steady-state amplitude and oscillation frequency vary by a small amount δA and $\delta\omega$. If δA and $\delta\omega$ decrease with time, the oscillator is considered to be stable about its operating point.

Using the Taylor series expansion around ω and A and neglecting the higher-order terms yields Eqs. (16) and (17) (Ref. 7):

$$-A\frac{\partial R_d}{\partial A}\delta A + A\frac{\partial(R_d - R_c)}{\partial\omega}\delta\omega = 0 \tag{16}$$

$$A\frac{\partial X_d}{\partial A}\delta A + A\frac{\partial(X_d - X_c)}{\partial\omega}\delta\omega = 0 \tag{17}$$

In order for the δA and $\delta\omega$ go to zero, the determinant of the above system of linear equations must not be equal to zero. Kurokawa's stability condition (8) is given as

$$\left.\frac{\partial R_d(A)}{\partial A}\right|_{A=A_0}\left.\frac{\partial R_c(\omega)}{\partial\omega}\right|_{\omega=\omega_0} - \left.\frac{\partial X_d(A)}{\partial A}\right|_{A=A_0}\left.\frac{\partial X_c(\omega)}{\partial\omega}\right|_{\omega=\omega_0} > 0 \tag{18}$$

For maximum stability the large signal device impedance and the circuit impedance should intersect at right angles at the operating point.

INJECTION LOCKING OF MICROWAVE OSCILLATORS

Injection locking of a microwave oscillator is accomplished by applying a small signal to the free running microwave oscillator provided that the frequency of the small signal is close enough to the free running frequency of the oscillator. Through injection locking a microwave oscillator to a low-noise low-power source, its phase noise can be improved. It has been shown that the phase noise of the injection locked oscillator close to the oscillation frequency can be reduced to that of the locking source (9). Furthermore, the microwave oscillator can also be frequency-modulated as long as the frequency of the injection signal stays within the locking range. By increasing the amplitude of the injection signal, the locking range can be increased.

The effect of the injection signal can be studied by adding a small voltage $v(t) = v\cos(\omega t + \phi)$ in series with the equivalent circuit for a microwave oscillator as shown in Fig. 4. In this case, Eqs. (16) and (17) should be modified as

$$-A_0\frac{\partial R_d}{\partial A}\delta A + A_0\frac{\partial(R_c - R_d)}{\partial\omega}\delta\omega = v\cos\phi \quad (19)$$

$$A_0\frac{\partial X_d}{\partial A}\delta A + A_0\frac{\partial(X_c + X_d)}{\partial\omega}\delta\omega = v\cos\phi \quad (20)$$

For a fixed-frequency oscillator near its oscillation frequency we can assume that the device negative resistance is not a function of frequency. Furthermore, we assume that the device reactance is also independent of the current amplitude. In this case the Eq. (20) can be rewritten as

$$A_0\frac{\partial(X_c + X_d)}{\partial\omega}\delta\omega = v\sin\phi \quad (21)$$

or

$$\delta\omega = \frac{v\sin\phi}{A_o\frac{\partial(X_c + X_d)}{\partial\omega}} \quad (22)$$

By applying Foster's reactance theorm for a series resonant circuit ($\partial X/\partial\omega = 2RQ/\omega$) to the circuit in Fig. 4, we obtain

$$Q = \omega\frac{\partial(X_d + X_c)}{\partial\omega} \quad (23)$$

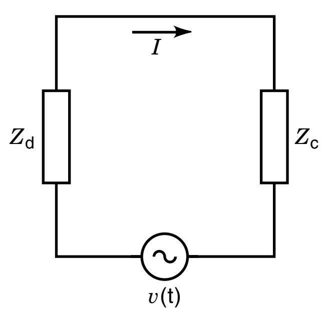

Figure 4. Simplified block diagram of an injection-locked oscillator. $V(t)$ is the injection signal.

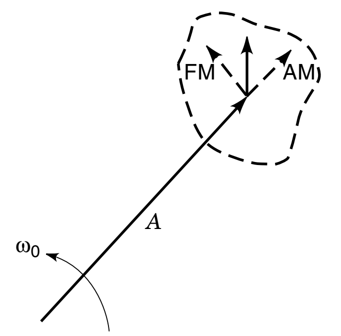

Figure 5. Vector representation of AM and FM noise in an oscillator.

Substituting Eq. (23) in Eq. (22) yields

$$\frac{\partial\omega}{\omega} = \frac{v\sin\phi}{2QA_0R_c} \quad (24)$$

The largest locking bandwidth is obtained when $\phi = \pi/2$. Then Eq. (24) can be written in the following form:

$$\left(\frac{\delta\omega}{\omega}\right)^2 = \frac{1}{4Q^2}\frac{P_i}{P_o} \quad (25)$$

where P_i and P_o are injection signal power and oscillator output power, respectively. Equation (25) shows that in order to increase the locking bandwidth, one can decrease the oscillator circuit Q and increase the injection signal power. It should be noted that Q is the external quality factor of the resonant circuit.

NOISE IN OSCILLATORS

The spectral purity of an oscillator is degraded due to the random fluctuations of its amplitude, frequency, and phase. Noise generated in the active device and passive components modulate the signal produced by the oscillator. This results in AM and FM and PM noise in oscillators. The oscillator noise in radio receivers and transmitters sets limits on their operation. The local oscillator phase noise in a receiver down converts the adjacent channels into intermediate frequency (IF) thereby limiting receivers immunity to nearby interference. In a radio transmitter the phase noise can overwhelm the nearby weak channels. In general the AM noise can be eliminated by using a limiter at the output of an oscillator. The vector representation of noise in an oscillator is shown in Fig. 5. The amplitude and angular frequency of the noise free oscillator are A and ω_o, respectively. The oscillator's noise is represented as a vector with random amplitude and angular frequency. The oscillator's amplitude, frequency, and phase fluctuations result in continuous sidebands around its operating frequency. The sources of random sideband noise in an oscillator include thermal noise, shot noise, and flicker ($1/f$) noise. The phase noise very close to the oscillation frequency is determined by flicker ($1/f$) noise of the active device. The basic model for phase noise in oscillators was described by Leeson (10). The FM noise spectrum for an oscillator is

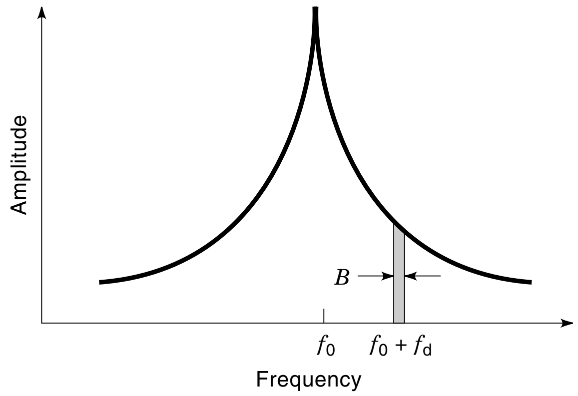

Figure 6. Spectrum of FM noise in an oscillator.

shown in Fig. 6. Normally the FM noise of an oscillator is measured as the ratio of noise power P_n in one sideband contained in a specified bandwidth at an offset frequency f_d compared to the total output power here referred to as P_o. Such power ratio is usually expressed in decibels below carrier (dBc). The single sideband FM noise at an offset frequency f_d is given by Eq. (26):

$$\left(\frac{P_n}{P_o}\right)_{SSB} = \frac{1}{2Q^2}\left(\frac{f_0}{f_d}\right)^2 \frac{KTBM}{P_o} \tag{26}$$

where K is Boltzmann's constant, T is the absolute temperature, B is the bandwidth in hertz, and M is the excess noise measure. The FM and PM noise can also be represented in terms of the rms frequency and phase deviation by Eq. (27) and Eq. (28), respectively:

$$\Delta f_{rms} = \frac{f_0}{Q}\sqrt{\frac{KTBM}{P_o}} \tag{27}$$

$$(\Delta\Phi_{rms})^2 = \left(\frac{\Delta f_{rms}}{f_d}\right)^2 \tag{28}$$

The relation between the single sideband phase noise (in dBc) and rms phase noise is given by

$$\text{single sideband phase noise (in dBc)} = 20\log\left(\frac{\Delta f_{rms}}{\sqrt{2}f_d}\right) \tag{29}$$

The AM noise of an oscillator can be expressed by Eq. (30) as

$$\left(\left[\frac{P_n}{P_o}\right]\right)_{AM,SSB} \frac{\dfrac{2KTBM}{P_o}}{\left(2Q\dfrac{f_d}{f_o}\right)^2 + S^2} \tag{30}$$

where S is the saturation factor. The value of S depends on the dependency of device negative resistance with the amplitude of oscillation. Normally, S^2 is a very large number compared to $4Q^2$ $(f_d/f_o)^2$ near the carrier. At relatively moderate frequency offsets, the AM noise power per hertz is much smaller than FM noise power per hertz (7).

The expression for the FM noise in injection-locked oscillators is given (11) as

$$(\Delta f_{rms})^2 = \frac{\dfrac{f_d^2 KTBM}{P_o}}{\left(Q\dfrac{f_d}{f_o}\right)^2 + \dfrac{P_i}{P_o}\cos^2\phi} \tag{31}$$

This expression indicates that the FM noise of an injection locked oscillator can be as small as that of the locking oscillator for small f_d.

GUNN OSCILLATORS

The Gunn diode (12) is a negative resistance device that operates based on transfer of electrons between two valleys in the conduction band of a semiconductor material. For this reason it is also referred to as a *transfer electron device* (TED). A Gunn device may be described as a slab of GaAs that exhibits negative resistance when dc bias is applied. In this respect the Gunn diode is not precisely a diode since it does not contain any junctions. Gunn diodes are capable of producing a few milliwatts to a few watts with efficiencies up to 15–20%. The operation frequency of Gunn diodes reaches up to 100 GHz. Gunn diodes have a good phase noise performance. The equivalent of the Gunn diode chip can be described by a parallel RC circuit where the value of R is negative. Gunn diodes exhibits a negative resistance at low frequencies as well as RF frequencies. Precautions must be made in order to avoid parasitic oscillations due to resonances introduced by the bias circuit. This can be achieved by making the bias circuit lossy at low frequencies in order to stabilize it.

The Gunn diode oscillator can be made in waveguide, coax, and microstrip lines. The most widely used Gunn oscillator circuit is an iris coupled waveguide cavity oscillator. This circuit has high stability and low phase noise due to its high external Q. Furthermore, one can mechanically or electronically tune the oscillation frequency. Figure 7 shows the iris-coupled waveguide cavity Gunn oscillator. The resonance frequency of the waveguide cavity is determined by the distance between the iris to the effective back wall and should be approximately a half-wavelength. The effective backwall is in between the diode's post and the physical back wall. If the diode's post is near the cavity's side wall, then the effective backwall is very close to the physical backwall. By adjusting

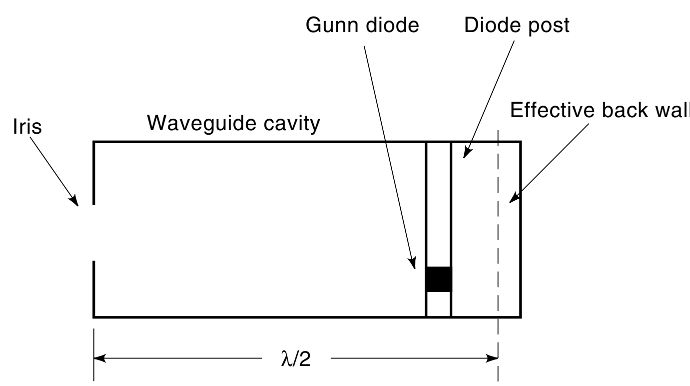

Figure 7. An iris-coupled waveguide cavity Gunn diode oscillator.

the diameter and the position of the diode's post with respect to sidewalls, optimum impedance match to the cavity resonator can be achieved. In order to mechanically tune the frequency of the cavity, a dielectric rod can be inserted into the cavity resonator. The electronic tuning of oscillation frequency can be accomplished by mounting a varactor diode inside the cavity using another post. By adjusting the varactor bias the oscillation frequency is tuned. This is a simple method to frequency-modulate such as oscillator.

IMPATT OSCILLATORS

The term IMPATT is an acronym for impact avalanche transit time. IMPATT diode is used in the design of solid-state microwave and millimeter-wave oscillators with operating range of frequencies from several gigahertz to above 200 GHz. IMPATT diode exhibits negative resistance due to 180-degree phase delay of the current with respect to the voltage. This phase delay is due to (1) the finite delay between the applied RF voltage and the current due to avalanche breakdown and (2) the subsequent transit of carriers through a drift region. One version of IMPATT operation was first proposed by W. T. Read in 1958 (13). The first demonstration of an IMPATT oscillator was reported in Ref. 14. The two most commonly used versions of IMPATT diodes are the single-drift region and double-drift region. IMPATT diodes produce 0.5 W to 4 W with efficiencies usually greater than 10% (up to 30 GHz) and 4% (up to 100 GHz). IMPATT diodes differ from Gunn diodes in several respects: (1) The output powers of IMPATTs are up to 10 times higher than those of Gunn diodes, (2) IMPATT diodes are more efficient than Gunn diodes, (3) Operating voltage for IMPATT diodes is higher than that of Gunn diodes (20 to 100 V for a X-band IMPATT as compared to 10 V or less for a X-band Gunn), (4) IMPATT diodes are noisier than Gunn diodes, and (5) the circuit design using IMPATT diodes is more difficult because the magnitude of their negative resistance is about one order of magnitude lower than that of Gunn diodes.

IMPATT diode is a current-controlled device and therefore it requires a current source for bias. The IMPATT diode equivalent circuit consists of a series RC circuit where R is a negative number. Since IMPATT diode is a current-controlled device, it is usually placed in a series resonant circuit for stable oscillators. Usually a radial line (Fig. 8a) or a coaxial line (Fig. 8b) quarter-wave transformer is used to match the low negative resistance of IMPATT to a waveguide cavity.

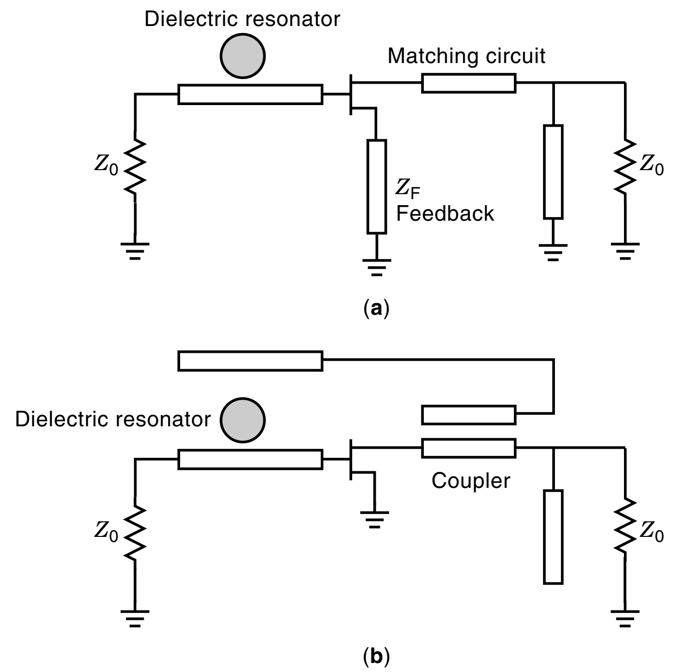

Figure 9. Dielectric resonator oscillator circuits using a) series feedback b) shunt feedback.

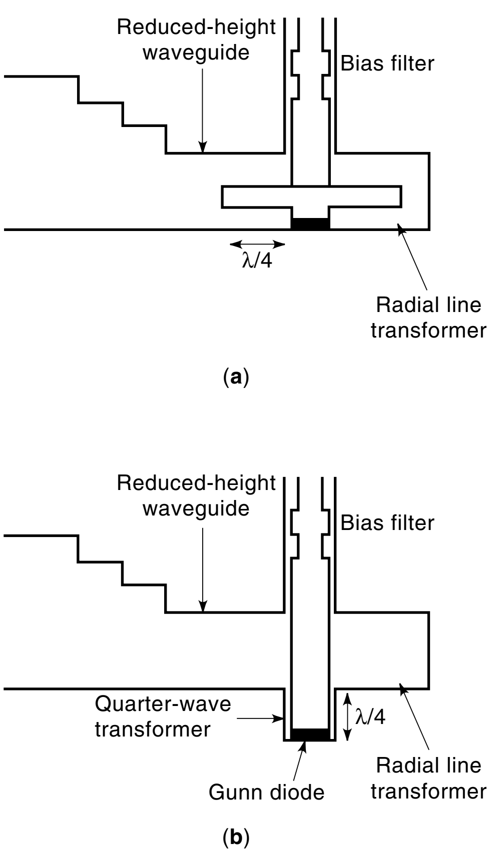

Figure 8. IMPATT diode oscillators in waveguide using (a) a radial line transformer and (b) a coaxial line transformer.

DIELECTRIC RESONATOR OSCILLATORS

Microstrip resonators have a limited unloaded Q on the order of 100. In order to reduce the phase noise of oscillators, it is necessary to increase the resonant circuit's external Q. A low cost technique to increase the resonator's external Q is to couple a dielectric resonator to a microstrip circuit. Typical dielectric resonators have a cylindrical geometry with a dielectric constant between 10 and 100. The $TE_{01\delta}$ of the cylindrical dielectric resonator can be coupled to a microstrip line by placing it on top of the substrate close to the microstrip line. The distance between the resonator and microstrip line determines the coupling factor. The unloaded Q of dielectric resonators is on the order of several thousand. The resonance frequency of a dielectric resonator can be adjusted with aid of a

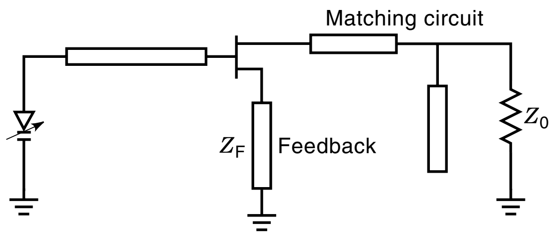

Figure 10. A varactor-tuned oscillator.

movable metal plate placed on top of the dielectric resonator. Figure 9 shows different configurations of dielectric resonator oscillators.

ELECTRONICALLY TUNABLE OSCILLATORS

In order to construct electronically tunable oscillators, either a varactor diode is connected to the resonator or a yttrium iron garnet (YIG) sphere is used to construct the resonator. In varactor tuned oscillators by changing the reverse bias voltage across the varactor, the varactor capacitance and thereby the resonance frequency can be tuned. The tuning range is dependent on the varactor's capacitance ratio C_{max}/C_{min}. This ratio is largest for hyper-abrupt varactor diodes. The frequency tuning limit of a varactor-tuned oscillator is determined by (4)

$$\frac{\omega_{\max}}{\omega_{\min}} = \sqrt{\frac{C_{\max}}{C_{\min}}} \tag{32}$$

A typical varactor-tuned oscillator circuit is shown in Fig. 10.

Another technique for construction of an electronically tunable oscillator is to make the resonator by using a magnetic material. Single-crystal yttrium iron garnet is a magnetic material that resonates at microwave frequencies when subjected to a dc magnetic field. The resonance frequency is directly proportional to the applied magnetic field; hence a linear frequency tuning over wide band can be achieved by adjusting its biasing dc magnetic field. The YIG resonator consists of an RF coupling loop, an electromagnet, and a YIG sphere. YIG is a low loss material and YIG resonators can provide unloaded Q factors as high as several thousand. YIG-tuned oscillators are commonly used in sweep generators. The frequency tuning range of the YIG oscillators is determined by the bandwidth over which the active device provides negative resistance. Figure 11 shows the diagram of a YIG-tuned oscillator. YIG-tuned oscillators with oscillation frequencies as high as 60 GHz have been reported (15).

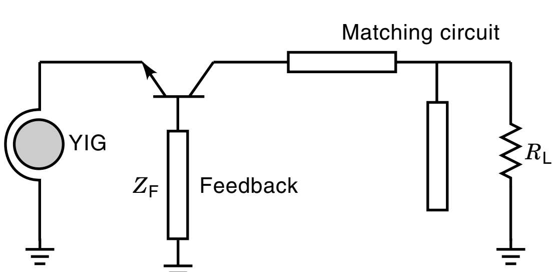

Figure 11. Diagram of a YIG-tuned oscillator.

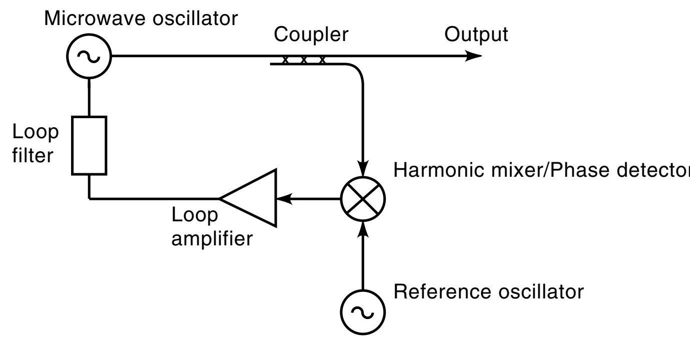

Figure 12. Block diagram of a phase-locked loop.

PHASE LOCK LOOPS

For many applications the frequency stability and phase noise of free running microwave oscillators such as Gunn and IMPATT oscillators are not adequate for system applications. By phase locking a free running microwave oscillator to a stable lower-frequency reference oscillator, its stability and noise performance can be improved. In general the free running oscillator takes on the stability and noise performance of the reference oscillator when phase-locked. The block diagram of a microwave phase-locked oscillator is shown in Fig. 12. For locked systems, a highly stable crystal oscillator operating at HF or VHF is used as a reference oscillator. A dc-coupled harmonic mixer/phase detector is used to mix the output of the crystal oscillator with that of the free running microwave source. If the difference in frequency is small, phase locking will take place. A search mechanism is generally used so that the loop will be forced to tune through a stable lock point if the initial difference frequency is too large for capture to occur.

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MICROWAVE OVENS. See MICROWAVE HEATING.

MICROWAVE PACKAGING. See PACKAGING RF DEVICES AND MODULES.