DC–AC POWER CONVERTERS

DEFINITIONS

Static converters are circuits using static valves that convert or control electrical energy. They enable the energy flow between different systems to be controlled. When ac and dc systems are coupled four basic functions are possible:

1. Rectification, the conversion of ac into dc
2. Inversion, the conversion of dc into ac
3. Dc conversion, the conversion of dc of a given voltage and polarity into that of another voltage and (where applicable) reversed polarity
4. Ac conversion, the conversion of ac of a given voltage, frequency, and number of phases into that of another voltage, frequency, and (where applicable) number of phases

These four basic functions in the conversion of electrical energy are performed by corresponding types of converters: the rectification function by a rectifier, the inversion function by an inverter, the dc conversion function by a dc converter, and the ac conversion function by an ac converter, the latter also referred to as a frequency converter.

In the case of rectifiers and inverters the energy direction is preset, but with dc and ac converters the direction of energy flow can change in general.

Figure 1 shows single-phase and three-phase dc–ac converters (inverters). In the inverter mode the energy flows from dc to ac.

TYPES OF INVERTERS

Taking three-phase bridge and single-phase bridge connections as examples, four different types of inverter are shown in Fig. 2. Circuits with external commutation (line or load commutation) can be designed with thyristors without turnoff

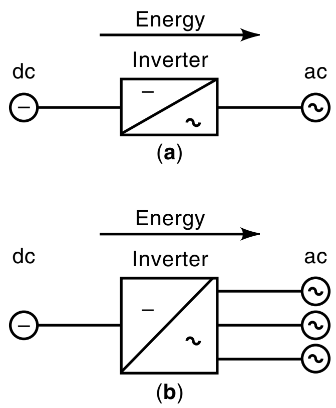


Figure 1. Dc–ac converter (inverter) converts energy from dc to ac: (a) single phase and (b) three phase.

arms. Circuits with self-commutation require turnoff valve arms, in other words, gate-turnoff semiconductor devices (1).

The two types of inverters with impressed current contain smoothing reactances on the dc side. The two with impressed voltages have smoothing capacitances on the dc side. Besides, they need antiparallel connected free-wheeling diodes. Most important in practice are the inverters with impressed current and external commutation and also the inverters with impressed voltage and self-commutation. The inverter type with impressed current and self-commutation is also (but less often) applied. In practice, the inverter type with impressed voltage and external commutation is used only with a series resonant circuit, but has been at least considered for use with synchronous machine (2,3) and for utility application (4).

In inverter operation mode, energy flows from the dc (current or voltage impressed) to the ac side. All circuit variants shown may also be operated in rectifier mode. Then electrical energy flows from the single- or three-phase ac side to the dc side.

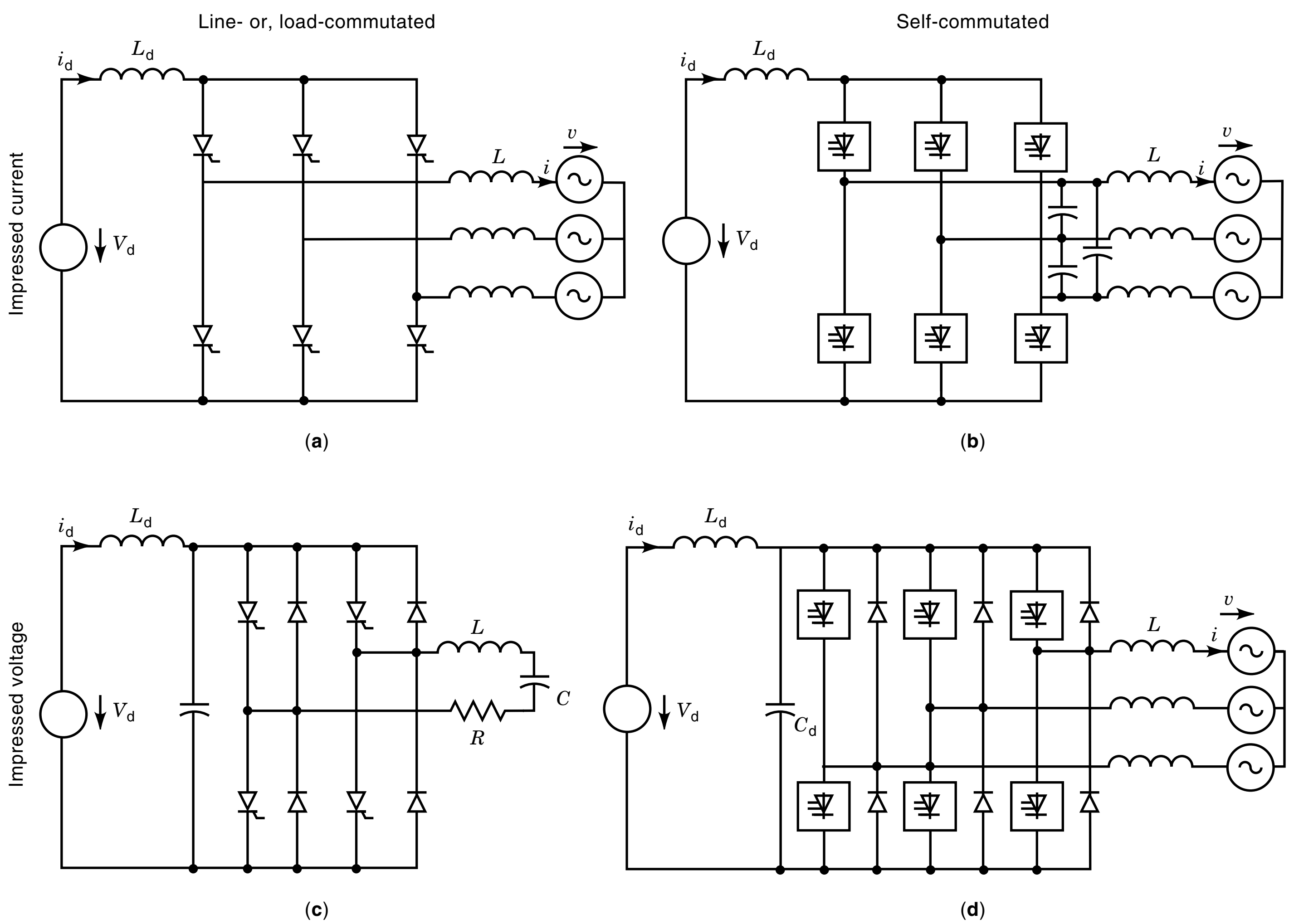


Figure 2. Basic types of inverters: (a) phase controlled rectifier or inverter, (b) current impressed self-commutated inverter, (c) series resonant inverter, (d) PWM-inverter; most significant inverters are (a) and (d).

Table 1. Development of Power Electronics

Era	Dominant Devices	Converter/Inverter Circuits	Engineering Solution
1900 to 1960	Commutator Periodic mechanical switches Mercury arc valves Selenium rectifiers Electronic tubes Magnetic amplifiers	Line-commutated rectifiers and inverters Frequency converters (cyclo-converters)	Rectifier Battery charger Feeding of electrolysis Speed control of dc machines HVDC
1960 to 1980	Silicon diodes Thyristors Bipolar transistors	Self-commutated choppers and inverters Load-commutated inverters	Dc–dc conversion Speed control of ac machines Uninterruptible power supply Inductive heating
1980 to 1990	Bipolar transistors MOSFETs GTOs IGBTs	PWM inverters and resonant circuits at higher frequency	Sinusoidal currents and voltages: Lineside and loadside reactive power compensation Active power filter
1990 to 2000	Improved power devices: Light-triggered MOS-controlled Thick-film technology Smart power HVIC Sensors		Digital process control Self-adjusting More speed-controlled ac drives Intelligent interaction with line and load

HISTORY OF DEVELOPMENT

Converter circuits (rectifier or inverter) work with electrical valves and have benefited continually from the improvement of semiconductor devices. Table 1 shows the development of power electronics from the very beginning with thyratrons and mercury arc rectifiers. The semiconductor era, beginning with silicon diodes and thyristors (about 1960), started a rapid development of available converter circuits and a steady increase in applications.

The first inverters were realized with periodically operated mechanical switches: the commutator of dc machines or mechanical polarity reverser and vibrating converters.

Line-commutated inverters were first achieved with thyratrons and mercury arc valves. Self-commutated inverters need quenchable arms, which were achieved by fast thyristors (inverter thyristors) and auxiliary arms (McMurray–Bedford inverter, McMurray inverter). Semiconductor devices capable of turnoff via gate impulses, such as bipolar transistors, metal–oxide–semiconductor field-effect transistors (MOSFETs), gate-turnoff thyristors (GTOs), and insulated-gate bipolar transistors (IGBTs) simplify the circuits, especially for pulse-width modulation (PWM) inverters and resonant inverters for high frequency.

INVERTER VALVES

The valves applied in dc–ac power converters (inverters) are nowadays exclusively semiconductor devices. With externally commutated inverters these are phase-control thyristors, which are available for repetitive blocking voltages up to 10 kV and for maximum average on-state currents up to 4 kA and higher. With phase-control thyristors inverters can be built ranging from the lower power level of several kilowatts up to the megawatt range.

Self-commutated inverters depend on turnoff devices. The first available gate-turnoff semiconductors were bipolar transistors, which however were limited with respect to their switching capability (hundreds of volts and tens of amperes).

Since 1980 more gate-turnoff power semiconductor elements have been developed and implemented, which have vastly promoted modern power electronics (5,6). Included here are improved bipolar transistors (high power, with fine structure, also with shorter switching times), field-effect transistors (MOSFETs), GTOs, and IGBTs. With these turnoff devices converters can be built for each power range. These components can be turned on and off by controlled gate currents or voltages.

Bipolar transistors and IGBTs are suitable for applications in the medium and lower power ranges, and GTOs and static induction thyristors (SITHs) for the upper power range. Because their excessive stored charge causes switching losses, bipolar devices are restricted to low to medium switching frequencies. This also applies to bipolar power transistors in a Darlington circuit. Gate-turnoff power devices are available (on the world market) for every power range (Table 2).

The switching behavior of power semiconductors is the main factor in determining their frequency range. MOSFETs and static induction transistors (SITs) are the fastest devices and are therefore suitable for the highest frequencies. Bipolar transistors have an upper frequency limit of several kilohertz. If a fine structure design is used, for example, in silicon ring emitter transistors (SIRETs), switching times can be considerably shortened and the frequency range raised to between 10 and 20 kHz. IGBTs also allow frequencies of up to 20 kHz. Non-punch-through (npt) design allows for voltage blocking

Table 2. Gate Turnoff Devices (Maximum Power, 1998)

Devices	V (V)	I (A)	t_{off} (μs)	P_{max} [a] (kV·A)	Frequency Range[b] (kHz)	Expenditure for Gate Driver
Bipolar transistor	1200	800	15 to 25	400	0.5 to 5	Medium to high
	550	1000	5 to 10	200		
SIRET	1000	80	1 to 3	30	2 to 20	High
IGBT	3300	1200	1 to 4	1200	1 to 10	Low
MOSFET	1000	28	0.3 to 0.5	10	5 to 100	Low
SIT	1400	25	0.1 to 0.3	15	30 to 300	Medium
GTO	6500	4000	10 to 25	5000	0.2 to 1	High
SITh	2000	600	2 to 4	300	1 to 10	High
MCT	3000	300	5 to 10	300	1 to 10	Low

[a] Maximum converter output power in three-phase bridge connection (one device per arm).
[b] PWM, hard switching.

capability up to 3.3 kV. Punch-through (pt) devices are suitable for even higher blocking voltages. Large GTO thyristors, on the other hand, are considerably slower. High-frequency (HF) GTOs with improved switching properties and SITHs allow higher switching frequencies.

In the case of GTO thyristors, the peak gate turnoff current is between $\frac{1}{3}$ and $\frac{1}{5}$ of the anode current. The turnoff delay times for large GTOs are around 10 μs or longer, GTOs in a special housing like hard-drive GTOs and integrated gate commutated thyristors (IGCT) can be operated with a peak gate turnoff current equal to the anode current (7). This shortens the delay time to below 1 μs and simplifies GTO series connection. In view of the switching losses incurred, their frequency is limited to below 1 kHz. In converters with load-independent voltage (choppers, inverters, and pulse-controlled inverters with capacitive storage), the switchable devices are always used in combination with inverse-parallel diodes. Fast switching diodes are required. Some devices have integrated diodes, and to some extent parasitic diodes as well.

Gate Driver and Snubber

The power requirements for triggering (particularly for generating the turnoff pulses) are a significant factor in the evaluation of switched power semiconductor devices. The triggering requirements are low for unipolar devices (MOSFETs and SITs). They are also low for IGBTs (voltage control), but high for bipolar transistors (or intermediate in a Darlington circuit) and GTOs in PWM mode. In none of the devices, however, do the drive requirements represent an obstacle to obtaining an economical circuit.

Investigations of actual use show relatively high expenditure for current-driven elements and small expenditure for voltage- or field-controlled elements, with the exception of IGBTs and MOSFETs.

In Fig. 3 the typical structure of a gate driver is given: high-frequency power supply, galvanic isolation, energy storage in capacitors (especially for turnoff pulses), and transmission of the information signal (on–off) via an optocoupler.

The necessity for and design of suppressor circuits are determined by the switching behavior of power semiconductor devices and the physical limits to critical voltage and current rises. Such circuits are also referred to as snubber circuits. In the case of MOSFETs and transistors, $(dv/dt)_{crit}$ is not specified. The device essentially determines the current and voltage curves depending on the gate drive.

For gate turnoff and static induction thyristors, in contrast, critical voltage and current changes must always be observed when turning the device on or off; otherwise it will be destroyed (excessive di/dt) or will arc through (dv/dt too high at turnoff). From this it follows that, in order to protect switched power semiconductors, GTOs and SITHs must have suppressor circuits, although these may be smaller in the case of a SITh (due to the higher values of dv/dt and di/dt). Transistors can basically be operated without a suppressor circuit, because the voltage and current changes during turnoff are sufficiently limited by the device itself. However, a surge voltage limiter is in fact frequently provided.

With GTO converters, a conventional R_aC_aD circuit is usually applied in which the energy ($\frac{1}{2}C_aV^2$) stored in the snubber capacitor C_a is converted to heat in resistor R_a when the GTO is turned on again. The power dissipation rises proportionally to the switching frequency, and in the case of high-power inverters this results in considerable losses. The resistors (R_a) are large components. The snubber losses are of the same order of magnitude as the on-state power losses of the devices, and reduce the efficiency. This is why variants of snubber circuits with intermediate storage have been developed to obtain a loss reduction of 50% or more. This is particularly advantageous for GTO inverters used in electrical traction.

EXTERNALLY COMMUTATED INVERTERS

Externally commutated converters (rectifiers and inverters) need a separate source of voltage, not belonging to the con-

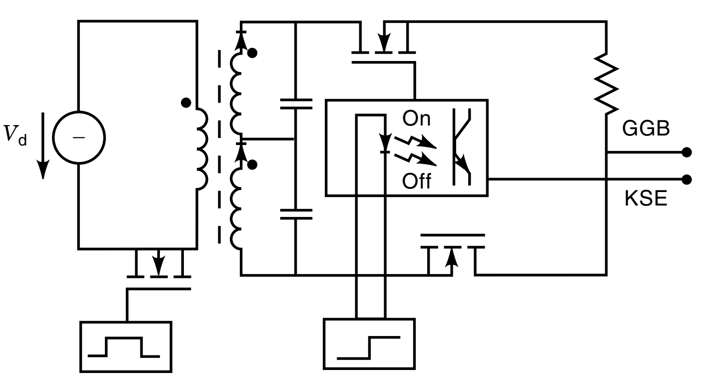


Figure 3. Typical gate driver: high-frequency intermediate circuit, galvanic isolation via transformer, on/off signal via photodiode.

verter, to provide it with the commutating voltage during the period of commutation. In the case of line-commutated inverters this source of ac voltage is the supply system; with load-commutated inverters it is the load (8).

Line-Commutated Inverter

Line-commutated inverters perform the basic functions of inversion and draw their commutating voltage from the single-phase or three-phase ac supply system, i.e., they use the voltages available in the supply system for commutation. The commutating voltage has the correct polarity during one half cycle only, i.e., the possible commutating zone in the case of converters with natural commutation is limited to this half cycle.

By definition the ideal no-load direct voltage V_{di} is the no-load direct voltage resulting from the phase voltage V on the secondary side of the converter transformer, ignoring resistive and inductive voltage drops. The commutating number q is the number of commutating events occurring during one cycle of the supply system within a group of mutually commutating valves. The pulse number p is defined as the total number of nonsimultaneous commutations of a converter connection during one cycle of the ac supply. Thus the ideal no-load direct voltage V_{di} is

$$\begin{aligned} V_{\mathrm{di}} &= \frac{s}{(2\pi)/q}\int_{-\frac{\pi}{q}}^{\frac{\pi}{q}} \sqrt{2}V\,\cos\omega t\,d\omega t \\ &= \frac{s}{(2\pi)/q}\sqrt{2}V\,\sin\omega t\Big|_{-\frac{\pi}{q}}^{\frac{\pi}{q}} \\ &= s\frac{q}{\pi}\sqrt{2}V\,\sin\left(\frac{\pi}{q}\right) \end{aligned} \tag{1}$$

The factor s is 1 for center tap connections and 2 for bridge connections.

Control Angle. With controlled converter valves, e.g. thyristors, transfer of the current to the next arm occurs only after triggering. Transfer can therefore be delayed with respect to the natural intersection of the phase voltages. The control angle α is defined as the time by which the instant of triggering is retarded from that with full modulation. The control angle is generally stated in electrical degrees. The converter is then said to have phase angle control.

The mean value of the dc voltage with a control angle α can be calculated by integrating within limits differing by the control

$$\begin{aligned} V_{\mathrm{di}\alpha} &= \frac{s}{(2\pi)/q}\int_{-\frac{\pi}{q}+\alpha}^{\frac{\pi}{q}+\alpha} \sqrt{2}V\,\cos\omega t\,d\omega t \\ &= \frac{s}{(2\pi)/q}\sqrt{2}V\,\sin\omega t\Big|_{-\frac{\pi}{q}+\alpha}^{\frac{\pi}{q}+\alpha} \\ &= s\frac{q}{\pi}\sqrt{2}V\,\sin\left(\frac{\pi}{q}\right)\cos\alpha \end{aligned} \tag{2}$$

From Eqs. (1) and (2) the important relationship for the ideal no-load direct voltage $V_{di\alpha}$ occurring at control angle α is obtained:

$$V_{\mathrm{di}\alpha} = V_{\mathrm{di}}\cos\alpha \tag{3}$$

which says that the mean value of the dc voltage of line-commutated converters varies as the cosine of the control angle α.

Operation in the Inverter Mode. The control angle α can be continuously increased from $\alpha = 0$ (maximum value of direct voltage V_{di}), whereby the output dc voltage varies in accordance with Eq. (3). At $\alpha = 90°$ the mean value of the dc voltage is zero. On further increasing the control angle beyond 90°, the mean value of the dc voltage becomes negative and continues to rise with negative polarity as the control angle is increased. At $\alpha = 180° - \gamma$ it reaches the maximum possible negative mean value. The extinction angle γ is the time, expressed in angular measure, between the moment when the current of the arm falls to zero and the moment when the arm is required to withstand steeply rising off-state voltage.

The range of control angles of $\alpha = 90°$ to $180° - \gamma$ with negative mean value of dc voltages is called operation in the inverter mode, because the direction of energy flow is opposite to that when operating in the rectifier mode.

Commutating Voltage. The commutation voltage v_k is a sinusoidal alternating voltage arising in a multiphase system as a result of the difference between the voltages of two mutually commutating phases. From the phasor diagram of the voltages v_1 and v_2 of two commutating phases separated by an angle $2\pi/q$ the commutating voltage can be calculated from

$$V_{\mathrm{k}} = 2V\,\sin\left(\frac{\pi}{q}\right) \tag{4}$$

In a three-phase system $V_k = \sqrt{3}V$ and is therefore equal to the line voltage.

Commutating Current. The waveform of current during a line commutation can be easily calculated. Assuming that the resistances R_k in the commutating circuit are neglected and commutating inductances L_k are the same size in each phase, the equation of the commutating voltage is given by

$$v_{\mathrm{k}} = 2L_{\mathrm{k}}\frac{di_k}{dt} \tag{5}$$

Here, i_k is the short-circuit current flowing in the commutating circuit. With a sinusoidal commutating voltage $v_k = \sqrt{2}V_k \sin \omega t$ one obtains from Eq. (5) the waveform of the short-circuit current i_k in the commutating circuit:

$$i_{\mathrm{k}} = \frac{1}{2L_{\mathrm{k}}}\int \sqrt{2}V_{\mathrm{k}}\,\sin\omega t\,dt \tag{6}$$

With the initial condition $t_0 = 0$, $i_k = 0$,

$$i_{\mathrm{k}} = \frac{\sqrt{2}V_{\mathrm{k}}}{2\omega L_{\mathrm{k}}}(1-\cos\omega t) \tag{7}$$

This equation reproduces the waveform of the short-circuit current, which is illustrated in Fig. 4. With phase angle control and a control angle α, the initial condition $\omega t_0 = \alpha$, $i_k = 0$ applies, i.e.,

$$i_k = \frac{\sqrt{2}V_k}{2\omega L_k}(\cos\alpha - \cos\omega t) \tag{8}$$

If the phase short circuit occurring when two valves are carrying current simultaneously were still to exist after the end of the commutation process, the short-circuit current would continue to rise. The maximum value with $\alpha = 0$ and $\omega t = \pi$ is:

$$2\sqrt{2}I_k = \frac{\sqrt{2}V_k}{\omega L_k} \tag{9}$$

In fact, in fault-free converter operation the short circuit is removed at the end of the commutating time because the current in the relieved valve becomes zero and the valve blocks. Depending upon the control angle α, the transient response of the commutating currents consists of the corresponding sections of the short-circuit current waveform. When $\alpha = 90°$, commutation occurs at the maximum value of the commutating voltage v_k with the greatest rate of rise; thereafter, the rate slows on further increasing the control angle α for operation in the inverter mode.

Overlap Time. The commutating time t_u, that is, the time span during which two commutating converter arms carry current simultaneously, is called the overlap time, and the corresponding overlap angle u is generally stated in electrical degrees. It can be calculated from Eq. (5) by integrating this equation over the commutating time:

$$\int^{t_u} v_k\,dt = \int^{t_u} 2L_k\frac{di_k}{dt}\,dt = \int^{t_u} 2L_k\,di_k \tag{10}$$

Since the commutating current i_k changes from zero to I_d during the overlap time t_u,

$$\int^{t_u} v_k\,dt = 2L_kI_d \tag{11}$$

With control angle α this becomes

$$\int_{\alpha/\omega}^{\alpha/\omega+t_u} \sqrt{2}V_k\ \sin\omega t\,dt = \frac{1}{\omega}\sqrt{2}V_k\left[-\cos\ \omega t|_{\alpha/\omega}^{\alpha/\omega+t_u}\right] = 2L_kI_d \tag{12}$$

producing

$$\cos(\alpha+u) = \cos\alpha - \frac{2\omega L_kI_d}{\sqrt{2}V_k} = \cos\alpha - \frac{I_d}{\sqrt{2}I_k} \tag{13}$$

The initial overlap angle u_0 with $\alpha = 0$ is obtained from

$$\cos u_0 = 1 - \frac{I_d}{\sqrt{2}I_k} \tag{14}$$

Figure 5 illustrates the waveform of voltage and current of a three-phase bridge connection at control angle $\alpha = 150°$, i.e.

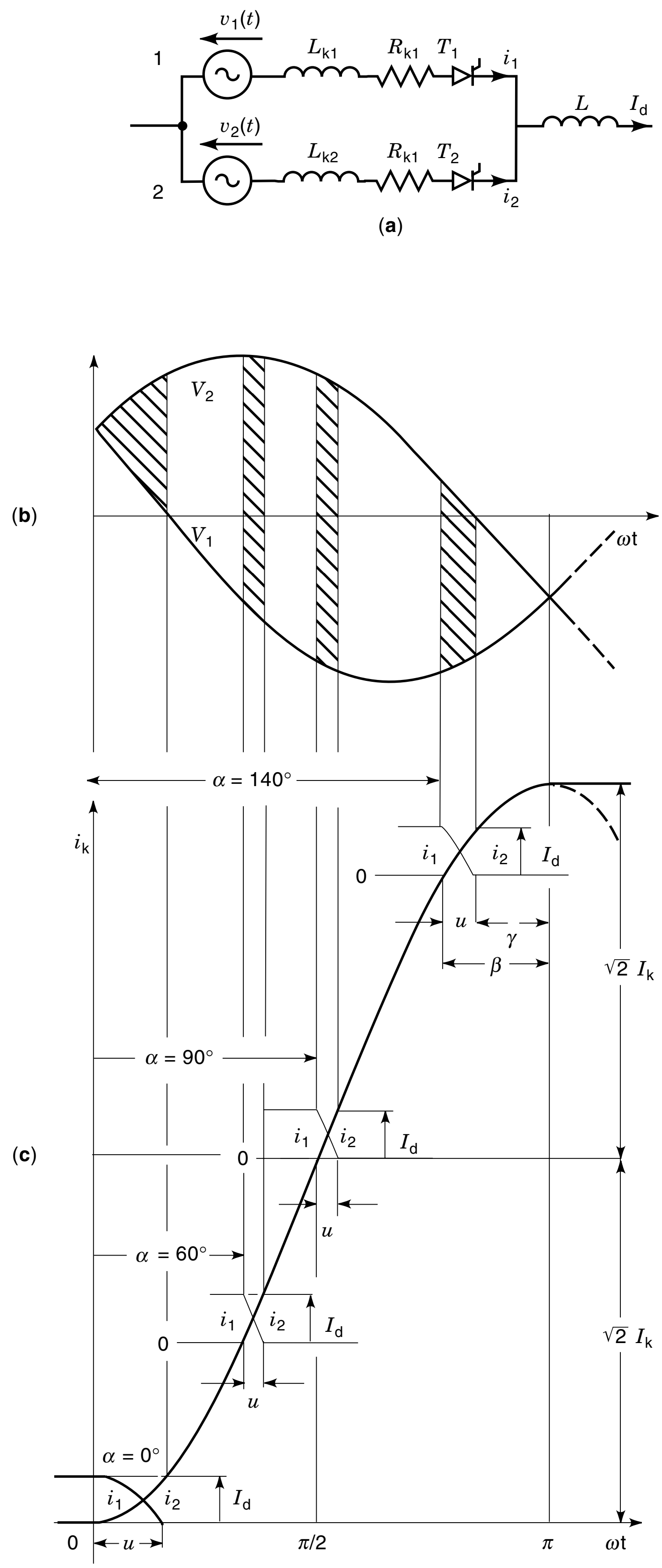


Figure 4. Line commutation: (a) commutation circuit, (b) commutation voltage, (c) commutation current; $\alpha \leq 90°$ rectifier operation, $\alpha \geq 90°$ inverter operation, γ = extinction angle.

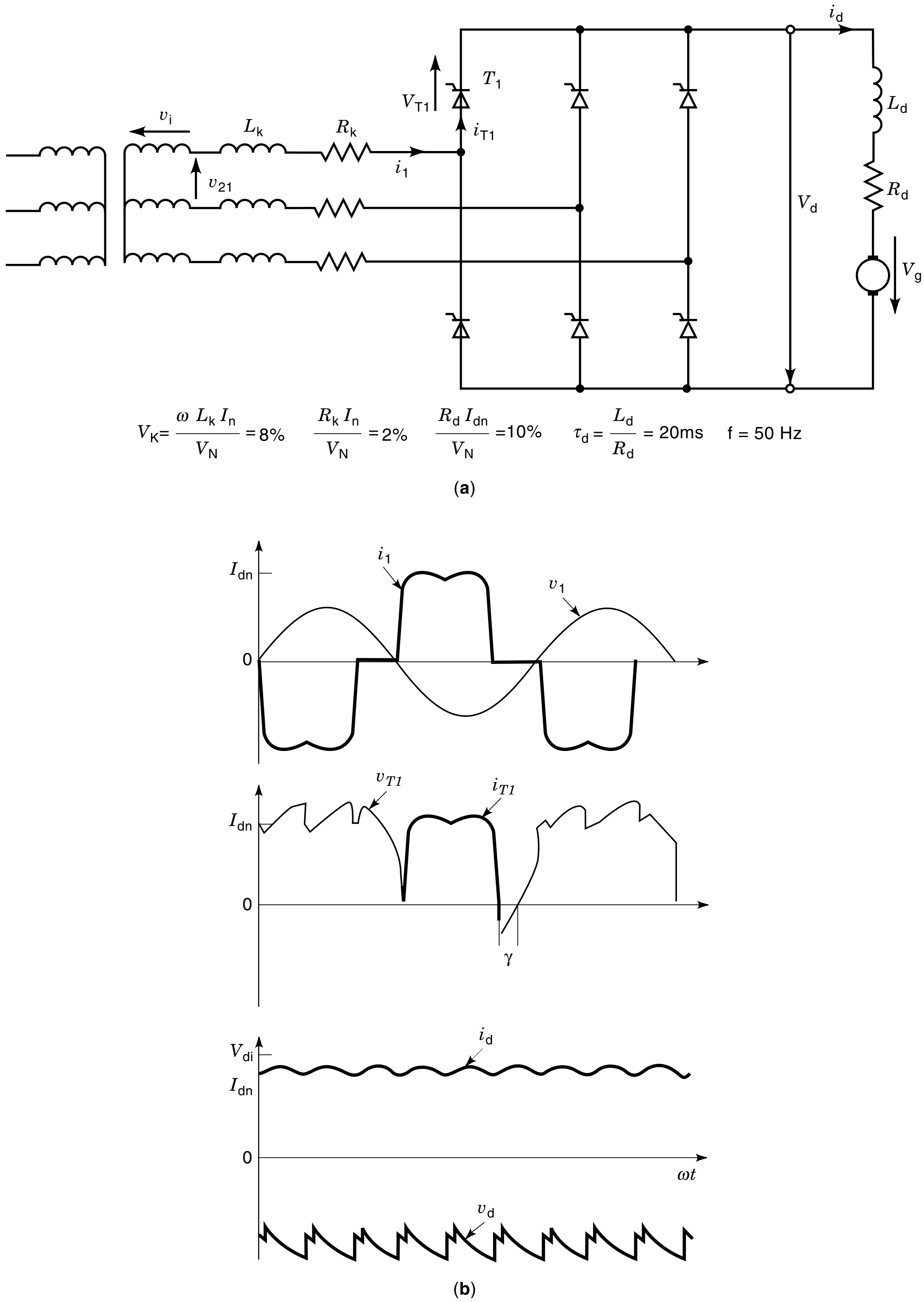


Figure 5. Line-commutated inverter: (a) three-phase circuit with given parameters, (b) voltage and current waveforms, $\alpha = 150°$; lineside current i_1 and voltage v_1 nearly in "antiphase"; extinction angle defined by negative voltage v_{T1}; also shown six pulse dc current i_d and dc voltage v_d.

operation in the inverter mode taking commutation into consideration. The resistances R_k as well as the inductances L_k in the commutating circuit are considered. It is assumed that there is a smoothing choke L_d on the dc side with ohmic losses.

During the overlap time t_u both commutating valves carry current. The commutating voltage v_k appears across the two commutating inductances L_k. When the inductances L_k are identical it is divided equally between them. The waveform of dc voltage v_d during the commutating time t_u is the average of the two commutating phase voltages. After completion of one commutation until the beginning of the next one, v_d is the phase voltage of the valve carrying the current. Due to the inductive voltage drop across the commutating inductances, the mean value V_d of the dc voltage is reduced.

Extinction Angle and Angle of Advance. The extinction angle γ necessary during operation in the inverter mode still has to be calculated, taking the commutation overlap time into consideration. In Fig. 5 operation in the inverter mode is illustrated with $\alpha = 150°$. From Fig. 4 the relationship

$$I_d = \sqrt{2} I_k[\cos\gamma - \cos(u+\gamma)] \tag{15}$$

can be derived, which can be converted into

$$\cos(u+\gamma) = \cos\gamma - \frac{I_d}{\sqrt{2} I_k} = \cos\gamma - 2d_x \tag{16}$$

where $d_x = \omega L_k I_d/\sqrt{2}\, V_k$.

The angle $u + \gamma$ is also designated the angle of advance β. Taking overloads I_d/I_{dN} and supply voltage dips V/V_N into account, the required value of β can be calculated from

$$\cos\beta = \cos(u+\gamma) = \cos\gamma - 2d_{xN}\frac{I_d}{I_{dN}}\frac{V_N}{V} \tag{17}$$

where $d_{xN} = \omega L_k I_{dN}/\sqrt{2}\, V_{kN}$.

During the operation in the inverter mode the extinction angle γ indicates the period of reverse blocking voltage across the valve. This is also called the holdoff time. The extinction angle γ must be larger than the circuit-commutated recovery time t_q of the inverter valves. If it falls below this value, a short circuit occurs between the relieving ac current phases and the dc side. This is called commutation failure.

Converter Connections. Converter connections are standardized. Two types of connections are distinguished: single-way and double-way connections. With single-way connections, the terminals on the ac side of the converter assembly and hence the converter transformer windings on the valve side (or, if a converter transformer is not present, then the connections of the ac system) carry a unidirectional current. They are each connected to only one principal arm. With double-way connections, the terminals on the ac side of the converter assembly and hence the converter transformer windings on the valve side (if there is no converter transformer, then the terminals of the ac system) carry an ac current without a dc component.

Center Tap Connections. Center tap connections are single-way connections. It is a feature of all center tap connections that the terminals of the converter arms with the same polarity are connected together and form one dc terminal while the second dc terminal is formed by the center tap of the ac system. Center tap connections have the code letter M. The commutation number q and pulse number p equal the number of converter arms.

Bridge Connections. Bridge connections are double-way connections. They comprise only arm pairs constructed from two converter principal arms at a center terminal. Each ac side connection is made at the center terminal of an arm pair. Terminals of common polarity of the arms pairs are connected together and thus constitute a dc terminal. Bridge connections have the code letter B. The commutation number q is equal to the number of ac terminals and hence to the number of arm pairs. Bridge connections having an even commutation number q have a pulse number p equal to q. If q is odd, then $p = 2q$.

Load-Commutated Inverters

Externally commutated converters that do not draw their commutating reactive power from the ac supply must draw it from the load.

With load-commutated inverters, the load provides the commutating voltage during the period of commutation. Since the converter constantly needs inductive reactive power for commutation, it is prerequisite for the operation of load-commutated converters that the load can provide this. The load current must, for this reason, have a capacitive component. This condition can be satisfied by an overexcited synchronous machine or by parallel and series resonant circuits.

Load-commutated inverters behave similarly to single-phase or three-phase line-commutated inverters. Basically the characteristics and equations derived for these converters also apply for load-commutated converters. If the load absorbs energy, the load-commutated converter operates in the inverter mode.

Motor-Commutated Inverter. Motor-commutated inverters are load-commutated inverters that draw their commutating reactive power from an appropriately magnetized synchronous machine serving as a load.

Their connection generally also renders reversal of the energy flow possible. The connection of motor-commutated inverters corresponds to that of line-commutated converters for rectification and inversion. By connecting a line-commutated rectifier and a motor-commutated inverter in series, an ac frequency converter is created with a synchronous machine as load (Fig. 6(a)). Generally an energy store (a smoothing inductance L_d) is provided in the dc link circuit, which decouples the instantaneous power of the converter I on the line side from the power of the converter II on the load side.

The converter I on the line side operates as a line-commutated rectifier when the connected synchronous machine is operating as a motor. The current I_d in the dc link circuit is smoothed by the inductance L_d.

The converter on the load side operates as a load-commutated inverter. If the direction of energy flow reverses, converter II must be modulated into operation in the rectifier mode and converter I into operation in the inverter mode. The dc voltages V_{dII} and V_{dI} thereby change their polarity. The direction of the current I_d is maintained. The synchronous machine then operates as a generator.

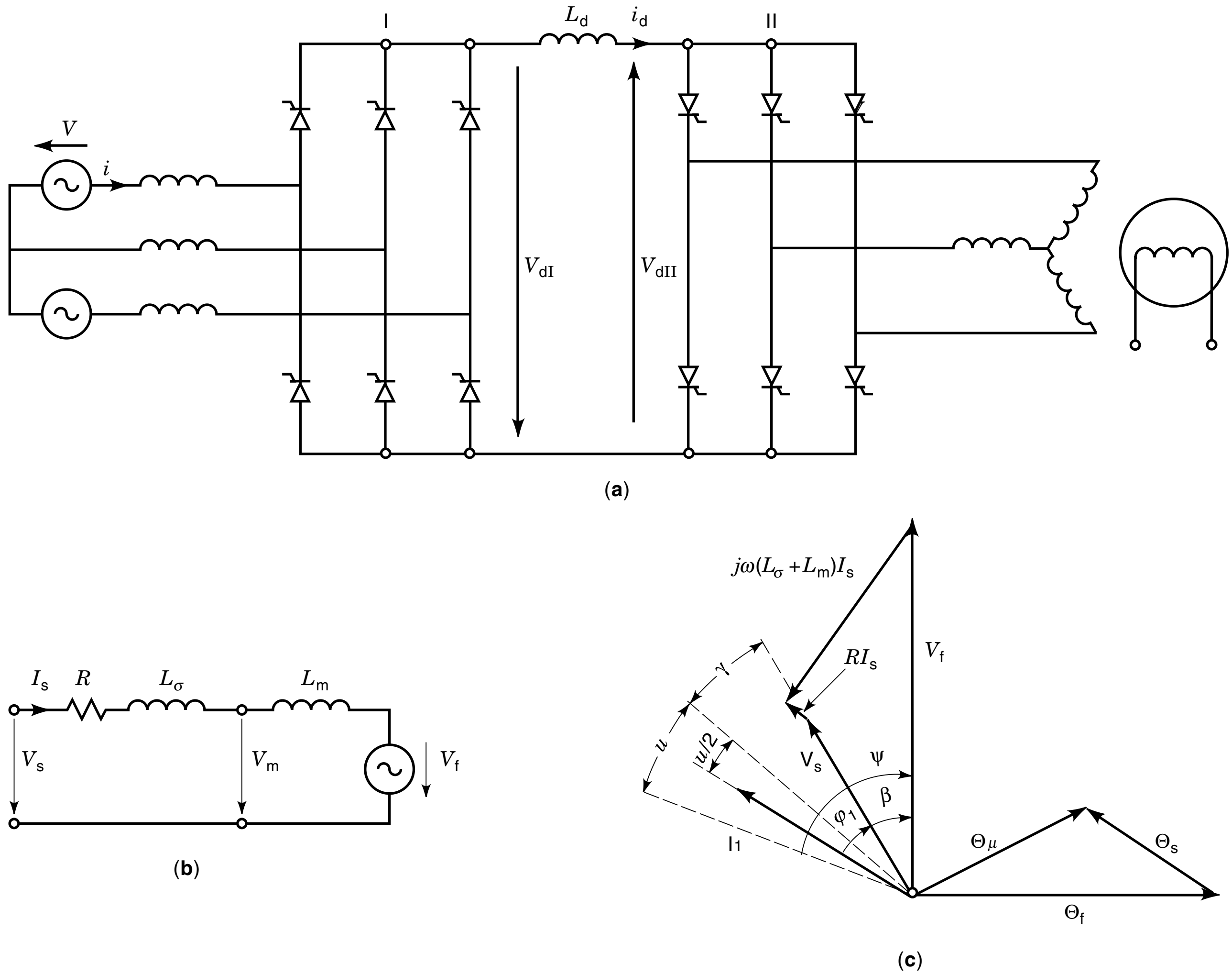


Figure 6. Motor-commutated inverter: (a) converter circuit, phase controlled converters on the line and on the motorside; synchronous machine mostly with brushless excitation, motor or generator operation possible, (b) single phase equivalent circuit, (c) phasor diagram, motor current I_s leads motor voltage V_s.

Figure 6(b) shows the per phase equivalent circuit of the synchronous machine. The synchronous machine load can only control converter II when the machine current has a capacitive component, i.e. when it leads the voltage. In Fig. 6(c) the phasor diagram of the synchronous machine is illustrated for this operating condition. V_s is the voltage of the machine, I_s the stator current. I_s leads V_s by the phase angle φ_1. Thus an overlap angle u produces the indicated extinction angle γ. This must be less than a certain minimum, called the holdoff interval necessary for the valves of converter II, as otherwise inverter failure would occur.

The extinction angle and overlap determine the power factor (for the fundamental) cos φ_1 of the synchronous machine. The phase angle is approximately $\varphi_1 \approx \gamma + u/2$. In order to keep the power factor high (but <1) despite changes in speed and torque, the holdoff interval (e.g. the extinction angle γ) can be controlled in a closed loop to a preset minimum value.

The other parameters shown in the phasor diagram are V_f, the voltage induced by the field of the rotor, and L_m, the effective machine inductance. Θ_f are the ampere–turns of the rotor, Θ_s the ampere–turns of the stator, and Θ_μ the resultant magnetizing ampere–turns. The angle β between V_s and V_f is the power angle of the synchronous machine; ψ is the so-called internal phase angle between the stator current I_s and the voltage V_p induced by the rotor. The torque developed by the synchronous machine is proportional to the cosine of this angle:

$$M \propto \Theta_s \Theta_f \cos\psi \tag{18}$$

With a given stator and field current, a synchronous machine therefore develops maximum torque when the internal phase angle ψ becomes zero.

At standstill of the synchronous machine, no commutating voltage is at first present on the secondary side, and special steps must be taken for starting. One possible starting method is to adjust the converter on the input side in synchronism with the low starting frequency, producing a discontinuous current in the dc link.

Parallel Resonant Circuit. A resistive–inductive load can be augmented by means of a capacitor into a parallel or series resonant circuit. The resonant frequency f_0 of the no-loss load circuit,

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \tag{19}$$

is called the rated frequency. The resonant frequency f_R of the freely oscillating load circuit subject to losses with the damping element δ is

$$f_R = \frac{\omega_0}{2\pi}\sqrt{1-\delta^2} \tag{20}$$

whereby

$$\delta = \frac{R}{2\omega_0 L} = \frac{R}{2}\cdot\omega_0\cdot C = \frac{R}{2}\sqrt{\frac{C}{L}} \tag{21}$$

These equations apply not only for a parallel but also for a series resonant circuit. The operating frequency f_B with which the resonant circuit inverter is operated is preset by the control system. In order to ensure a capacitive current component, the operating frequency of the parallel resonant circuit must lie above the resonant frequency. With the series resonant circuit it must lie below the resonant frequency.

In each arm of the parallel resonant circuit inverter (Fig. 7), there is a controlled converter valve whose current has an approximately rectangular waveform. Since the parallel resonant circuit does not permit sudden voltage changes, the inverter needs a smoothing inductance L_d on the dc side. The voltage v_2 of the load side is approximately sinusoidal. The current thereby transfers in direct commutation from one converter valve into the following one. The load current has a rectangular waveform and leads the load voltage by the phase angle φ. This is necessary to ensure the extinction angle γ.

The sinusoidal curve of load voltage v_2, which is identical to the capacitor voltage v_c, can be calculated from the energy balance between the dc and load sides. Taking into account only the fundamental voltage of v_2, one gets for a half cycle

$$V_d I_d \frac{T}{2} = I_d \int_{-\gamma/\omega}^{(\pi-\gamma)/\omega} \hat{v}_2 \sin\omega t\, dt \tag{22}$$

From this the peak value $\hat{v}_2$ of the load voltage can be calculated as a function of the dc voltage V_d and the extinction angle γ:

$$\hat{v}_2 = \sqrt{2}V_2 = \frac{\pi}{2\cos\gamma}V_d \tag{23}$$

As Eq. (23) shows, an increase in the extinction angle at constant output dc voltage V_d results in a voltage increase. For this reason the power output to the load circuit can be altered only to a limited degree by controlling the angle of advance $\beta = u + \gamma$. The output power can be controlled to a great extent by adjusting the dc voltage V_d.

Series Resonant Circuit. With the series resonant circuit inverter the resistive–inductive load L and R is augmented by a series capacitor C into a series resonant circuit (Fig. 8). In each arm of the inverter lies a controllable converter valve to which an uncontrollable diode is connected in antiparallel, so that the current can be carried in both directions.

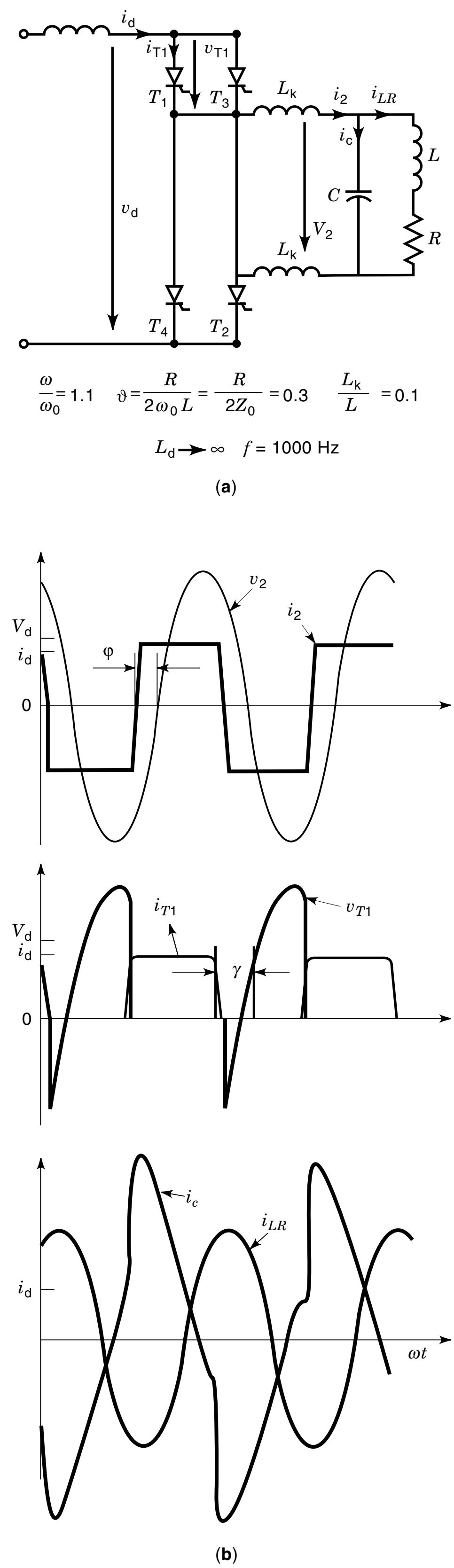


Figure 7. Parallel resonant circuit inverter: (a) single-phase bridge connection with given parameters and (b) voltage and current waveforms.

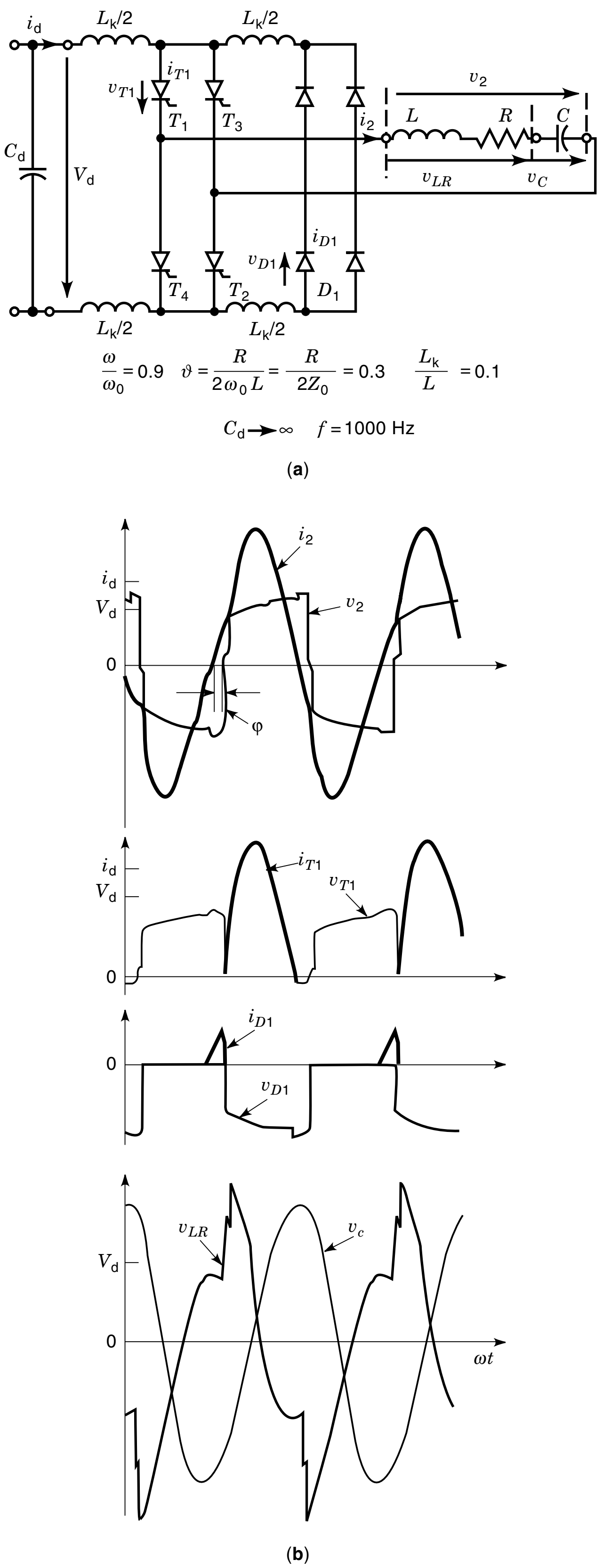


Figure 8. Series resonant circuit inverter: (a) single-phase bridge connection with given parameters and (b) voltage and current waveforms.

The series resonant circuit forces an approximately sinusoidal load current i_2, which is controlled alternately by the thyristors and the antiparallel diodes. The load voltage v_2, and hence also the valve voltage v_A, has an approximately rectangular waveform. The current commutates from the uncontrollable valve to the corresponding controllable valve connected in antiparallel. The load current leads the load voltage by the phase angle φ. This is necessary to maintain the required extinction angle γ.

As the series resonant circuit does not permit sudden current changes, as stable an output voltage as possible is needed while the thyristors are carrying the current energy flows into the load circuit and while the diodes are carrying the current energy flows back into the dc circuit. The output power can be controlled by altering either the extinction angle γ or the input direct voltage V_d. The fundamental of the load current i_2 can be calculated in the same way as with the parallel resonant circuit inverter from the energy balance between the dc and ac sides during a half cycle:

$$V_d I_d \frac{T}{2} = V_d \int_{\gamma/\omega}^{(\pi+\gamma)/\omega} \hat{\imath}_2 \sin \omega t \, dt \tag{24}$$

The peak value $\hat{\imath}_2$ of the load current is obtained from

$$\hat{\imath}_2 = \frac{\pi}{2 \cos \gamma} I_d \tag{25}$$

The attainable upper frequency limit of load-commutated resonant circuit inverters is determined mainly by the circuit-commutated recovery time of the thyristors. For the oscillations of the load to build up, particularly in the case of parallel resonant circuit inverters, a starting device is required to provide the commutating voltage needed for the first commutations after switching on the load-commutated inverter. For this, capacitive energy stores on the load and dc side are precharged.

Applications

Line-commutated inverters are used in speed-controlled dc motor drives when a reversing and regenerative braking mode is needed. A main application at very high power levels (up to the gigawatt range) is high-voltage dc transmission (HVDC), where one station is operated in the inverter mode (on the side of energy consumption) and is located far away from the second station. There are also some zero-distance asynchronous HVDC ties, such as for interconnecting 50 Hz and 60 Hz systems in Japan.

Motor-commutated inverters are applied for speed controlled synchronous motor drives in the medium or higher power range. Normally brushless synchronous machines with ac or rotating-field excitation and rotating diodes on the rotor are used (because of their low maintenance requirements). Parallel and series resonant inverters with load commutation are applied in the output frequency range of several hundred hertz (series resonant) up to more than 10 kHz (parallel resonant) for induction heating, hardening, and melting. See Table 3.

SELF-COMMUTATED INVERTERS

Self-commutated inverters fulfill the basic function of inversion, i.e. the conversion of direct current into alternating cur-

Table 3. Main Applications and Power Range of Inverters

Commutation	Impressed Current/Voltage	Circuit	Semiconductor Device	Main Application	Power Range
Line commutated		Phase controlled rectifier/ inverter	Phase control thyristors	Dc drives	kW to >10 MW
				HVDC	50 MW to >500 MW
Load commutated	Impressed current	Motor commutated inverter	Phase control thyristor	Speed controllable synchronous machine	500 kW to >10 MW
		Parallel resonant inverter	Fast thyristor	Inductive heating	
Load or self-commutated			MOSFET	Hardening	2 kW to >1 MW
		Series resonant inverter	SIT	Welding	
				Power supply	W to 1 MW
	Impressed voltage	Square wave inverter	MOSFET	UPS	W to 1 MW
		PWM inverter	IGBT	Speed controlled drives	kW to >10 MW
		Three-level inverter	GTO	Electric traction	>10 kW to >10 MW
			IGCT	Power factor correction	100 kVA to >100 MVA
Self-commutated				Active power filter	10 kVA to >1 MVA
	Impressed current	Capacitor commutated inverter	Thyristor	Single motor speed controlled drives (induction motor)	100 kW to >10 MW
		PWM inverter	GTO		
			IGCT		

rent. Since with these inverters commutation is carried out by energy stores belonging to the converter (quenching capacitors) or by turning off the converter valves via gate pulses, they are not dependent upon a separate source of alternating voltage such as an ac power supply system or the load.

The ac voltage generated can therefore generally be varied over a wide range of frequency. Under certain conditions control of the single-phase or multiphase ac output voltage is possible in addition. With multiphase self-commutated inverters a multiphase ac voltage system of variable frequency can be generated.

Parallel Inverter

Research into inverters started as early as the 1920s, when Alexanderson and Prince in the United States announced the so-called parallel inverter, in which two alternately working thyratrons were turned off by a capacitor (9,10).

Figure 9 shows stages in the development of inverters with self-commutation. Figure 9(a) shows the parallel inverter mentioned before, which is provided with a smoothing reactor on the dc side and is actually not capable of feeding inductive loads, since commutation occurs on the ac side. In 1932 Petersen for the first time described the so-called feedback diodes, which permit the supply of a lagging current (11). A turnoff of the controllable valves in this circuit with reverse-current diodes by means of a parallel capacitor requires the provision of commutating reactors for preventing the capacitor energy from being drained through the parallel-connected free-wheeling diodes. Particularly favorable conditions are obtained with a center-tapped commutating reactor as mentioned in Ref. 11, since there is an unimpeded current commutation from the controllable valve to the parallel diode.

In 1936 Tröger (12) described a circuit where, instead of a parallel capacitor between the phases, each controllable valve is provided with its own capacitor turnoff arrangement. The instant of turnoff of each valve can thus be freely selected. In 1938 Tröger also suggested (13) the connection of the free-wheeling diodes to a transformer tap, whereby circulating currents that may develop in the commutating reactor and the antiparallel valves are suppressed, [Fig. 9(b)].

This state of development of self-commutated inverters was already reached in the era of thyratrons and mercury-arc rectifiers. A technical realization, however, was possible only after the introduction of the thyristors, which feature a number of favorable properties for circuits with forced commutation, such as their short turnoff time. In 1961 McMurray and Shattuck (14) published an inverter circuit [Fig. 9(c)] with thyristors that is similar to the circuit with mercury-arc rectifiers previously suggested by Tröger. This circuit and variants of it were used in bridge arrangements for single-phase thyristor inverters up to megawatt ranges. The McMurray inverter circuit [Fig. 9(d)], especially, became widely applied with thyristor inverters (15).

Voltage-Source Inverter

Thyristor Inverter. The parallel inverter as illustrated in Fig. 9(a) is a self-commutated inverter in center tap connection with which the quenching capacitor C_k is arranged in parallel between the controllable valves, which relieve each other in carrying the current. The ideal output ac voltage is square-shaped. Its amplitude is $V_i = 2V_d$ (with a turns ratio $w_1/w_2 = 1$). The root-mean-square (rms) value V_{1i} of the fundamental oscillation is obtained from

$$V_{1i} = \frac{4\sqrt{2}}{\pi} V_d \tag{26}$$

With a negative countervoltage on the load side, the quenching capacitor C_k can lose part of its voltage after successful recharging to $2V_d$. In order to prevent this, blocking diodes are connected in series with the thyristors behind which the quenching capacitor is connected.

Single-Phase Bridge Connection. A single-phase self-commutated inverter can also be constructed in bridge connection. With this arrangement of the quenching capacitors it is a parallel inverter in two-pulse bridge connection. The reverse-

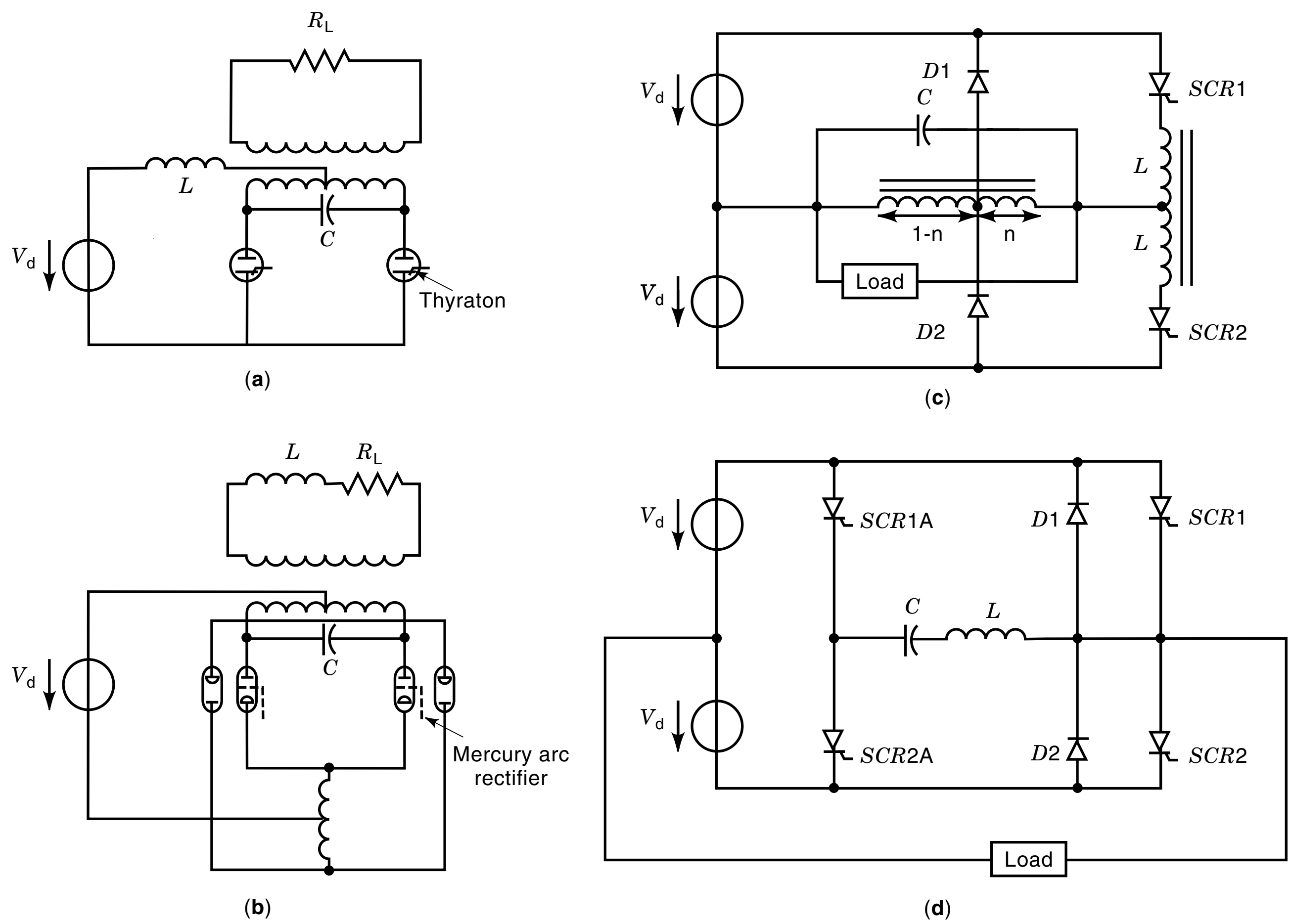


Figure 9. Self-commutated inverter: (a) parallel inverter with capacitor commutation, (b) parallel inverter with tapped transformer connection of feedback diodes, (c) McMurray–Bedford inverter, and (d) auxiliary impulse-commutated inverter (McMurray inverter).

current diodes again render alternating current of any phase possible.

The voltage generated on the ac side is also square-shaped. With constant direct voltage V_d the ideal output ac voltage is also constant. With the rms value $V_i = V_d$ the rms value V_{1i} of the fundamental oscillation is

$$V_{1i} = \frac{2\sqrt{2}}{\pi} V_d \tag{27}$$

Three-Phase ac Bridge Connection. Three-phase ac voltage systems can be created by self-commutated inverters in multiphase connections. Figure 10(a) shows the basic circuit of a three-phase self-commutated inverter. It is a six-pulse bridge connection with turnoff and regenerative arms. Under the assumption of 180° turn-on time of the devices $T_1 \ldots T_6$, such an inverter supplies the ac voltage indicated in Fig. 10(b) and 10(c). With a resistive load only the thyristor arms carry current. If reactive power occurs on the load side, the diodes (regenerative arms) also periodically participate in carrying the current. On reversal of the energy direction the diodes take over carrying the current.

At constant dc voltage V_d the output ac voltage is also constant with multiphase inverters, unless special steps (such as phase angle control or pulse control) are taken.

For the three-phase ac bridge connection the rms value V_i of the ideal line voltage v_{12} (120° square wave) is

$$V_i = \sqrt{\frac{2}{3}} V_d \tag{28}$$

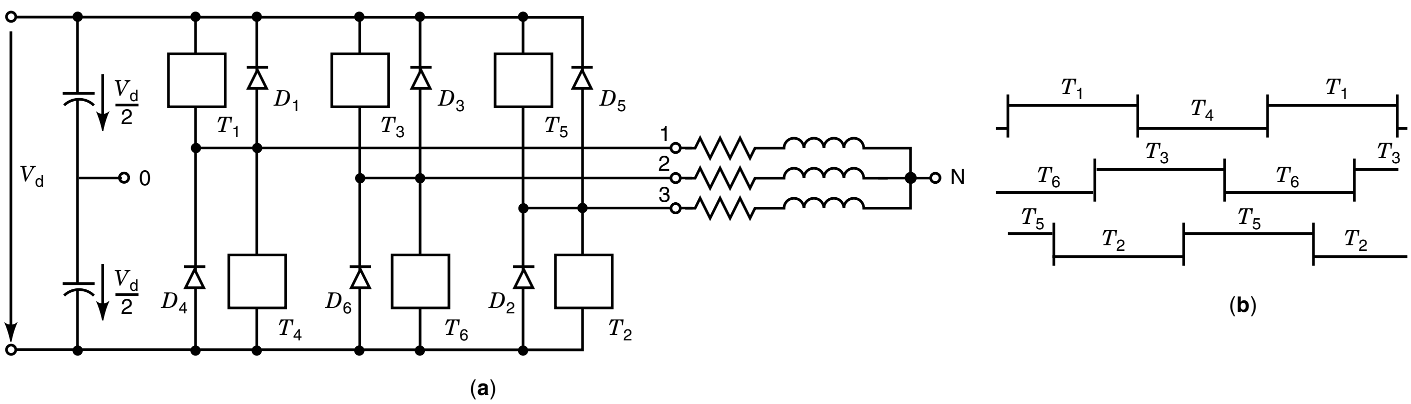
and for the fundamental oscillation

$$V_{1i} = \frac{\sqrt{6}}{\pi} V_d \tag{29}$$

With thyristor inverters the necessary quenching capacitors can be arranged in different ways. In combination with auxiliary thyristors and reactances, individual quenching of each inverter arm or of each inverter phase, or one capacitor for all inverter arms, has been applied. Modern self-commutated inverters use gate-turnoff semiconductor devices, e.g. transistors, MOSFETs, IGBTs, and GTOs.

Frequency Control. The frequency of the self-commutated inverters on the ac output side can be freely preset via the control circuit. In principle it can be continuously varied from zero to an upper limiting value.

With thyristor inverters this is determined by the holdoff interval necessary for the thyristors (circuit turnoff time t_q



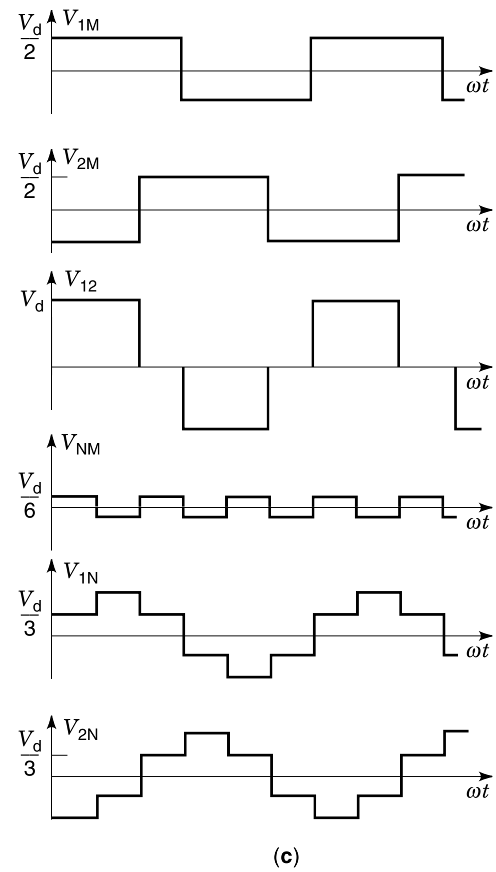


Figure 10. Voltage-impressed inverter: (a) inverter in three-phase bridge connection, turn-off device, and antiparallel diode in each branch; (b) scheme of turn-on time of the devices T_1 . . . T_6; and (c) voltage waveforms.

about 20 μs to 60 μs for inverter thyristors). Another limit is imposed by the losses produced by the capacitor quenching, which increase as the frequency.

With inverters equipped with gate-turnoff devices the upper frequency is limited by the switching characteristics of the semiconductor devices (see Table 2) and the switching losses (turn-on and turnoff). MOSFETs allow for the highest frequency (>200 kHz), IGBTs for medium frequency (<50 kHz), and large GTOs for low frequency (<1 kHz).

Voltage Control. Self-commutated inverters generate a single-phase or three-phase voltage system of adjustable frequency. According to Eqs. (26), (27), and (29) the ac output voltage is proportional to the dc voltage V_d. If the ac output voltage is also to be variable, the voltage control can be carried out either on the dc side in the inverter itself or on the ac side.

On the ac side the voltage can, of course, be altered by means of a regulating transformer. Its size grows, however, as the frequency becomes smaller.

Phase Control and Phase-Shifting Technique. With constant dc voltage V_d in the link circuit, voltage control can be carried out in the inverter itself. In control using the phase-shifting technique, the ac voltages of two uncontrollable inverters are phase-shifted by an angle α and added; see Fig. 11.

Shortening the voltage blocks causes the amplitude of the fundamental of the ac output to be reduced. However, at the same time the harmonics in the output voltage grow relative to the fundamental. For this reason these voltage control

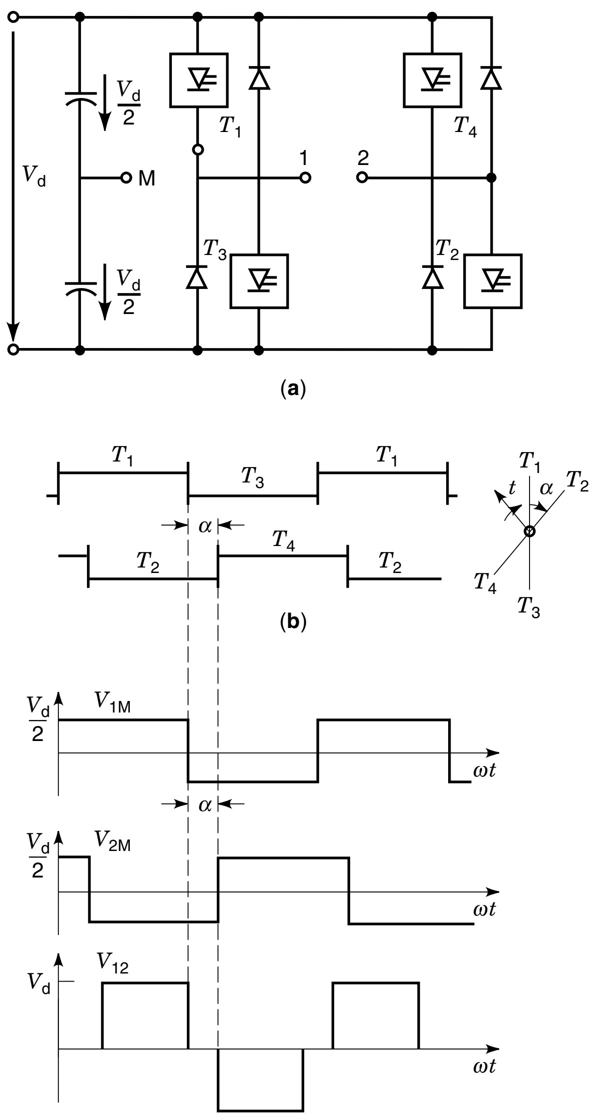


Figure 11. Voltage control by phase shifting (master and slave): (a) single-phase bridge connection, (b) scheme of turn-on time of the devices $T_1 \ldots T_4$, and (c) voltage waveforms.

techniques can only be employed within a limited range. Multiple-step polyphase inverters reduce harmonics in output voltage by phase-shifting in special transformer connections (12-pulse and higher).

Pulse Width Modulation. In control according to the pulse technique the converter arms are triggered and quenched several times in each period of the fundamental frequency. The pulse technique produces a sequence of individual conduction and idle intervals in the converter arm, the ratio of which determines the output voltage.

In Fig. 12 various pulse technique for voltage control with self-commutated inverters are reproduced. Depending upon the connection, either two voltage levels $+V_d$ and $-V_d$ or three voltage levels $+V_d$, 0, and $-V_d$ are possible. The mean value of the voltage over a half cycle can be controlled by altering the control factor $\lambda = T_1/(T_1 + T_2)$. Pulse techniques with three voltage levels have the advantage that the energy does not pulse unnecessarily between the load and the dc voltage link circuit.

If operation is not with a constant control factor λ, but the duration of the applied voltage blocks is matched to the waveform of the sinusoidal voltage reference, a good approximation to the sinusoidal fundamental is obtained. The fundamental frequency of the output voltage thus generated is said to be subharmonic. Besides the fundamental frequency, only harmonics of the chosen pulse frequency f_p and still higher harmonics occur at the load. Due to the inductance on the load side, a good approximation to a sine wave is obtained for the current waveform (16).

The pulse technique can also be extended to a direct two-step control of the load current. The load current then fluctuates within a preset current interval Δi by the (generally si-

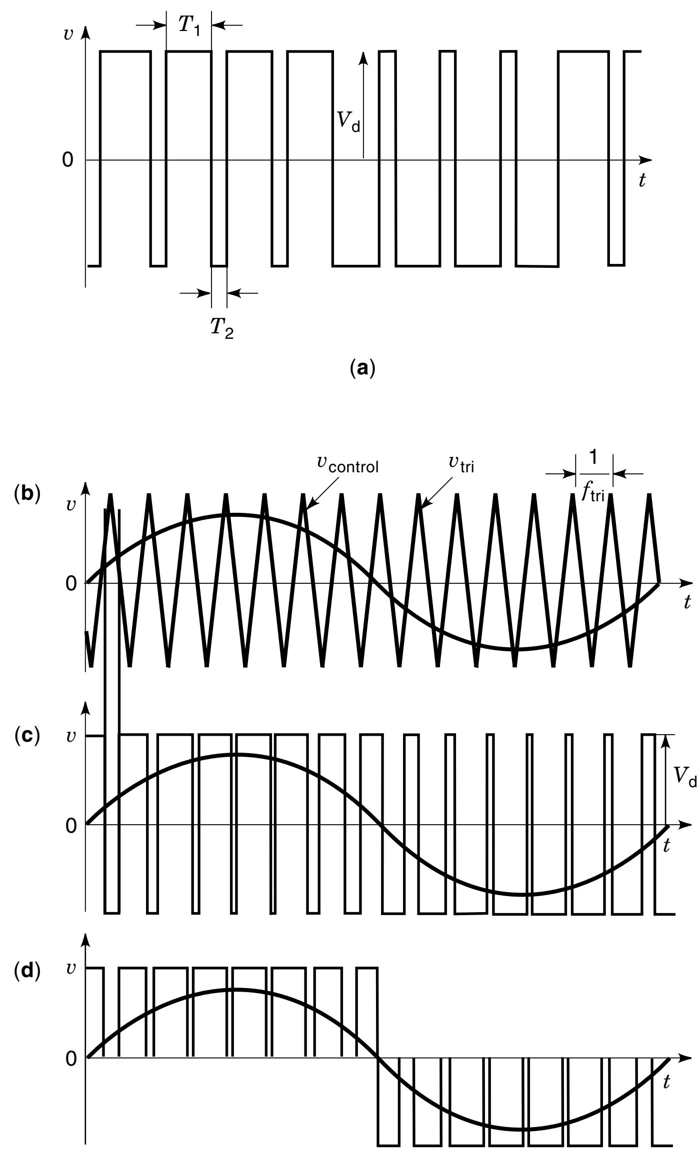


Figure 12. Pulse width modulation: $T_1/(T_1 + T_2)$ = const, two voltage levels; (b) comparison between sine wave and triangle control voltage; (c) T_1/T_2 variable according to sine function, two voltage levels; (d) three voltage levels.

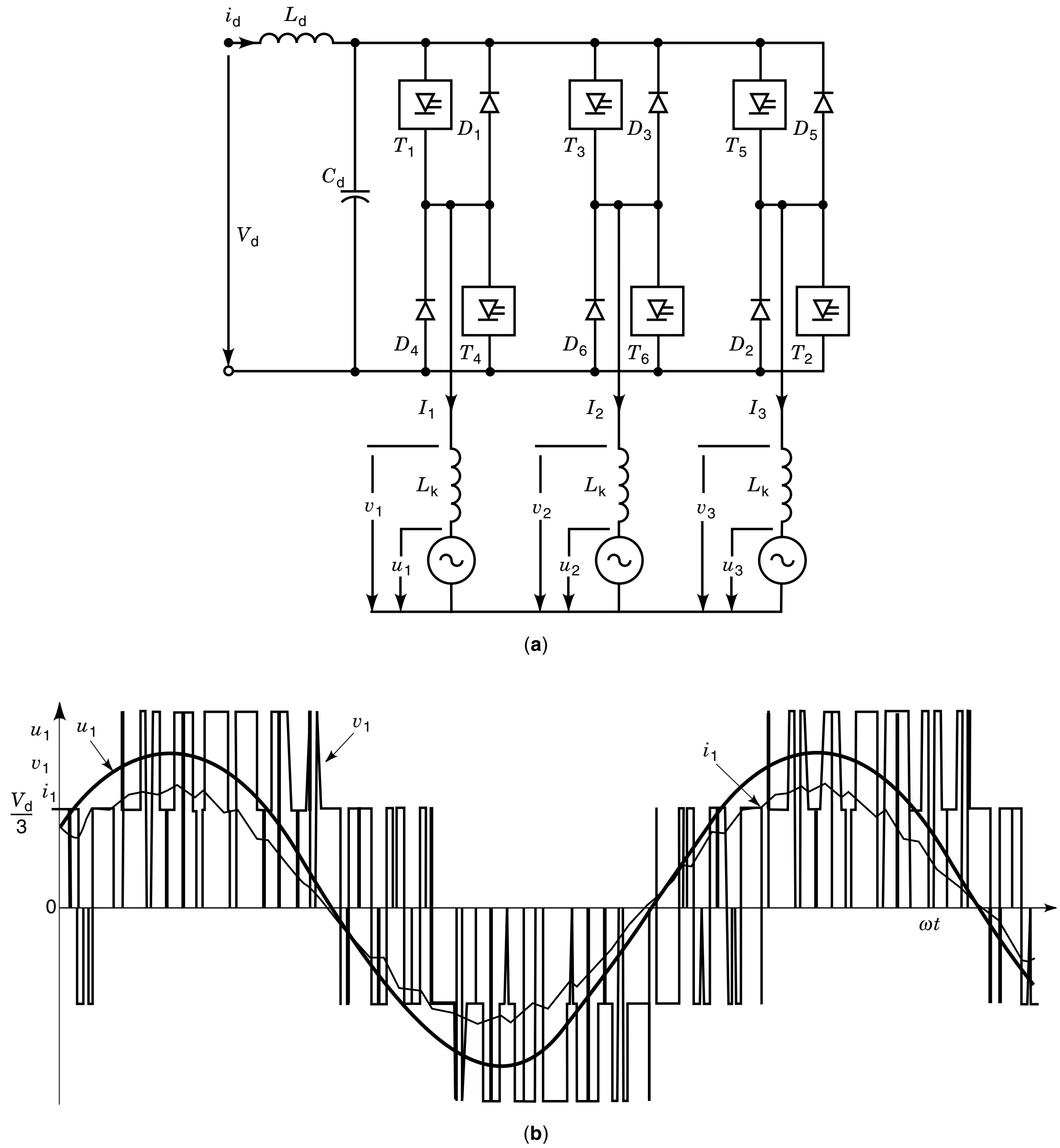


Figure 13. (a) PWM inverter in three-phase bridge connection and (b) voltage and current waveforms.

nusoidal) preset current reference (17). Suitable PWM methods have been developed, with respect to selected harmonic elimination and other features (18–20).

Pulse-Width-Modulated Inverter. A self-commutated inverter whose output voltage or current is controlled in open or closed loop according to the pulse techniques is called a pulse-width-modulated (PWM) inverter. With this inverter the total number of nonsimultaneous commutations during a cycle is increased (without increasing the number of semiconductor switches) by repeated turning on and off at the pulse frequency f_p. This increase can be used to reduce current and voltage harmonics, because it corresponds to an increase in the pulse number. With line-commutated converters an increase in the pulse number is only possible by means of a corresponding increase in the number of converter arms.

Single-Phase Bridge Connection. The coupling of a dc voltage source V_d with an ac voltage source can be achieved via a pulse-controlled inverter in single-phase bridge connection. Basically this connection corresponds to that of a dc power controller extended to four-quadrant operation. Each arm of the bridge consists of a quenchable converter valve and an antiparallel diode. The current i in the ac voltage source is continuously adjustable as long as the condition $V_d > \sqrt{2}\,V$ is satisfied. The current harmonics are determined by the inductance L_k on the ac side. Therefore, the inductance must be above a certain minimum value. since the power in a single-phase ac voltage source v pulses at double the fundamental

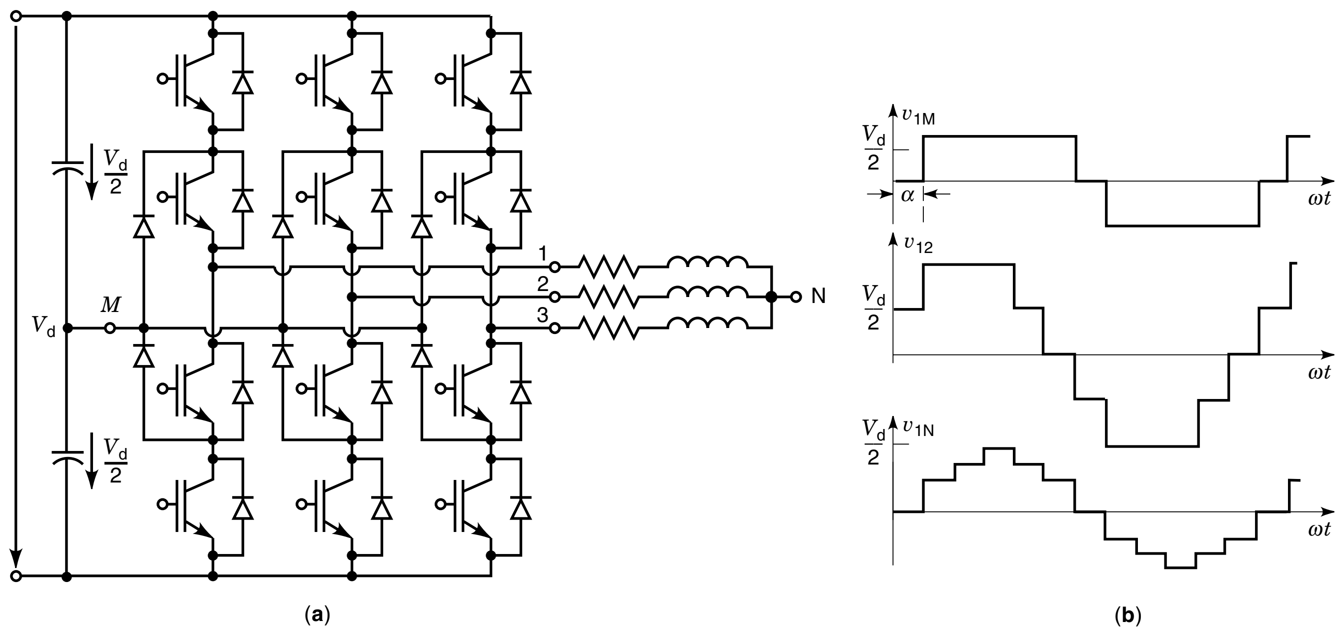


Figure 14. Three level inverter: (a) neutral point clamped PWM inverter and (b) voltage waveform with block control mode.

frequency, a sinusoidal current component of double frequency is superimposed upon the dc current I_d. This can be supplied by a resonant circuit tuned to this frequency. If this resonant circuit is omitted, a high-voltage ripple (double the fundamental frequency) will occur on the dc-link voltage.

Three-Phase Bridge Connection. Figure 13 shows the connection and the voltage and current waveforms of a pulse-controlled inverter in three-phase bridge connection. Each of the six arms of the bridge again consists of the antiparallel connection of a quenchable converter valve and a diode. Here again, the currents i_1, i_2, and i_3 are controlled to be sinusoidal. With a sinusoidal current waveform the sum of the phase powers drawn on the ac side is constant, i.e., equal to the power supplied by the dc voltage source V_d.

Multilevel Inverter. In order to increase the power output, voltage-source inverters can have three or more dc voltage levels created by separate smoothing capacitors. For each voltage level, bridge-connected turnoff semiconductor elements are arranged in combination with separating diodes. The voltages on the ac side then have five or more levels, which lead to lower harmonics or lower PWM frequency. Three-level inverters are especially applied with IGBTs in order to increase power output and dc voltage without putting individual devices directly in series (21,22). See Fig. 14.

Comparison of Losses of Turnoff Semiconductor Devices. Power semiconductors can be classified as unipolar or bipolar, according to their primary feature, the charge carriers involved in the transport of current. In unipolar devices (MOSFET, SIT), only one type of charge carrier is involved in current transport. High blocking voltages and a high current-carrying capacity therefore cannot be attained in one device. However, they can be controlled with very small amounts of energy (field-controlled) and—thanks to their short switching times—are suitable for very high switching frequencies. MOSFETs and SITs are restricted to the lower power range.

In the case of bipolar devices (diode, thyristor, bipolar transistor, IGBT, GTO, MOS controlled thyristor [MCT]), charge carriers are injected at a forward-biased *pn* junction from the heavily doped zone (emitter) to the lightly doped zone. The conductivity of the lightly doped zone can be raised in the on state by several powers of ten (conductivity modulation), leading to feasible devices with both high blocking and high current-carrying capabilities.

The losses in self-commutated inverters are determined by the semiconductor losses. The total losses in a semiconductor device consist of on-state power losses plus switching losses. In addition, there may be losses in the suppressor (snubber) circuit. On the other hand, the losses incurred when the device is in the off state are generally negligible.

The total losses are fundamentally dependent on the switching frequency in the case of bipolar and unipolar devices. As a result of their high forward voltage and resistance, the on-state losses for unipolar devices are high. Thanks to low switching losses, however, the total losses rise only moderately with switching frequency. The opposite is true of bipolar devices, which are better in the low-frequency range but whose losses climb rapidly with rising switching frequency due to the high levels of switching energy involved. IGBTs feature a good compromise, because they combine a relatively low forward voltage with moderate switching losses.

Figure 15 shows transistor losses in PWM inverters: Fig. 15(a) in principle, and Fig. 15(b) as measured in 5 kVA PWM inverters at various carrier frequencies. Only the IGBT allows for low losses at switching frequencies above the audio range.

Resonant Inverters. Resonant inverters are becoming more important in power electronics because the switching stress of power semiconductor devices is very low compared with that in PWM inverters. For the power semiconductor devices the on-state stress is quite different from that in normal hard-

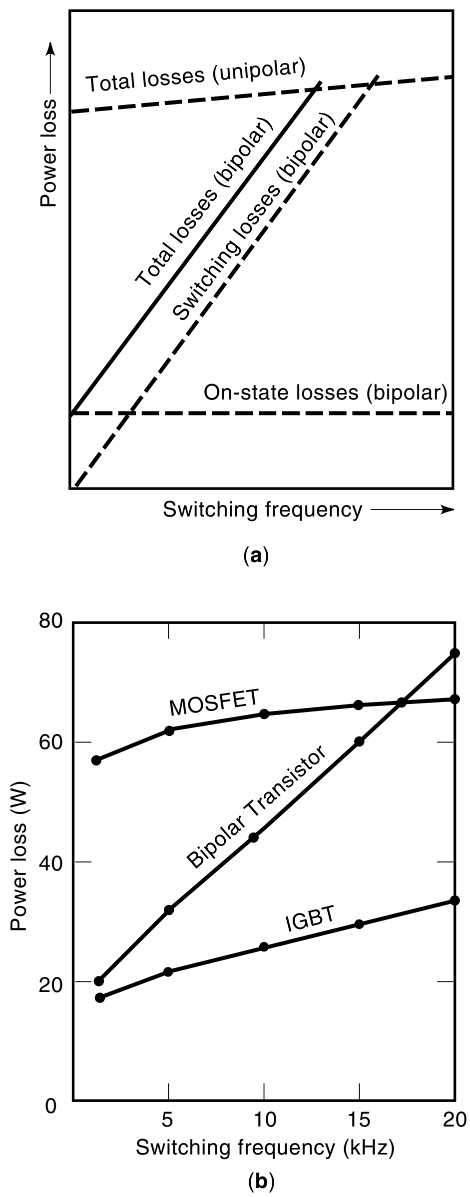


Figure 15. Losses of bipolar and unipolar power transistors: (a) principle losses and (b) losses in one transistor (as measured in a 5 kWA-PWM inverter).

switched PWM inverters, because higher peak values of device current and/or device voltage then occur. In soft-switched inverter applications these high peak device stresses can be avoided. Soft-switched inverter circuits seem to be a good choice for the medium and high power range. Resonant circuits allow higher switching frequencies. Reduced size of transformers, filters, and storage elements can be achieved. Capacitors and magnetic material for high frequencies must be developed.

Several circuit variants are under investigation. A detailed comparison of the qualities of these devices with those of normal hard-switched PWM converters (23,24) must be made.

The switching behavior in resonance converters is fundamentally different from that in hard-switched circuits, because either the voltage or the current is switched in the vicinity of the zero crossing. Evaluation of measurements in test circuits shows that the switching losses are negligible for MOSFETs in comparison with the on-state power losses. Similar evaluation of IGBT at up to 80 kHz shows that the switching losses are distinctly higher in this case than the low on-state power losses, and that they increase rapidly with frequency. Nevertheless, their efficiency, hypothetically calculated from a test circuit, is more than 90%.

Current-Source Inverter

Figure 16 shows a current-source inverter circuit for speed control of ac machines in industrial and other applications. The circuit is economic and suitable for high-power speed controlled single-motor drives. An inverter with impressed current and PWM with GTOs must be capable of blocking forward and reverse voltages. This circuit has rarely been applied so far (for high-performance elevators in Japan only).

Voltage and Current Harmonics

PWM of inverters reduces the voltage and current harmonics of low order. The pulse frequency ranges from 1 kHz up to over 1 MHz (with MOSFETs in the lower power range). For speed control of asynchronous and synchronous machines in the lower power range, meanwhile, standard PWM converters are used where the switching frequency lies between 1 kHz and 20 kHz. PWM inverters are also used for uninterruptible power supply (UPS) equipment. For big industrial drives and

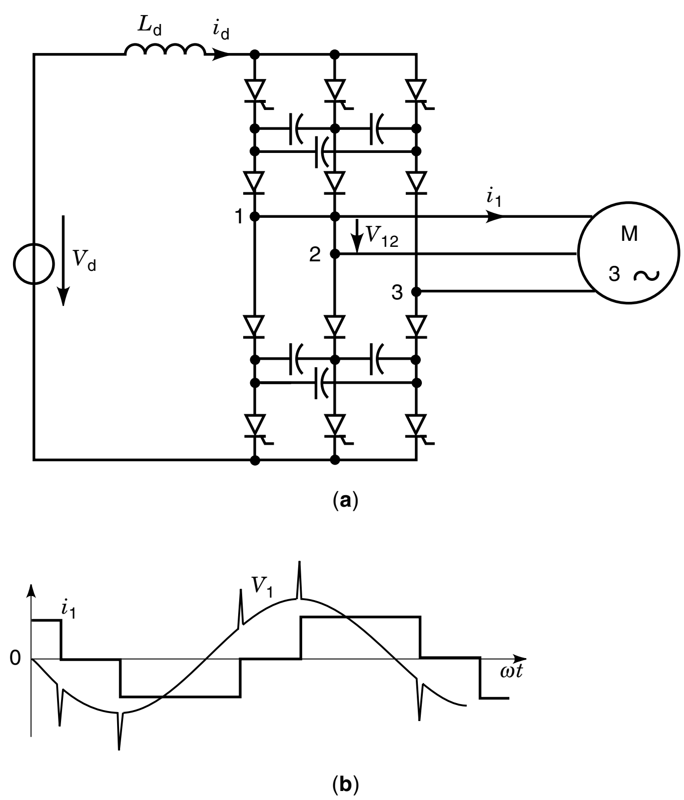


Figure 16. Capacitor commutated inverter, current impressed: (a) inverter circuit with induction motor, (b) voltage and current waveforms; the motor current is 120° square shaped, the motor voltage is sinusoidal with superimposed voltage spikes caused by capacitor commutation.

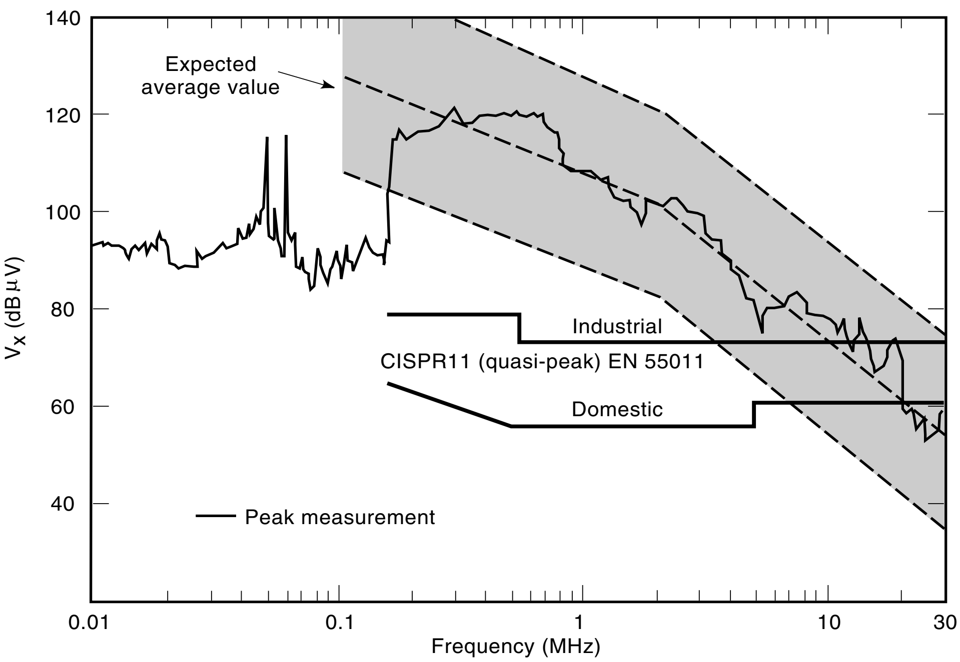


Figure 17. EMC of self-commutated PWM inverter (voltage impressed), measured noise spectrum of line conducted disturbances in the frequency range from 10 kHz to 30 MHz.

for electrical trains with asynchronous motors, PWM inverters with GTOs in the megawatt range are built. The pulse frequency for these is only a few hundred hertz.

The advantage of PWM in dc–ac converters (inverters) is the generation of sinusoidal currents and voltages load-side and also lineside as PWM ac–dc rectifiers if requested. At carrier frequencies above 16 kHz no audible noise is generated.

Electromagnetic Compatibility

The fast switching events, especially with MOSFETs and IGBTs, generate high dv/dt, which has to be limited by line-side and load-side filters.

Recently the upper limits specified for the noise spectrum generated by inverters in the frequency range from 10 kHz to 30 MHz were severely lowered. Figure 17 shows a measured noise spectrum of a hard-switching PWM inverter. The shaded band represents the noise level of inverters on the market in 1997. This band is 20 dB to 40 dB higher than the CISPR 11 limitation.

Special measures in the design of inverters, additional capacitors, and modified gate drivers are needed to fulfill the limits.

CONSTRUCTION AND DESIGN

Intelligent Power Modules

Power modules combine two or more semiconductor devices in one common case. A whole single- or three-pulse bridge connection can be encapsulated in one module.

The dissipation of a power module is limited by its thermal resistance, which in turn depends on the mechanical pressure applied. For economic reasons, plastic packages are preferred. Hence high voltages (>2 kV) present a special challenge. As the current per unit is limited, the possibility of connecting modules in parallel is important. Snubberless operation is a desirable economic feature. Therefore, low-inductance design is essential, leading to stripline techniques and flat cases. Present soldering techniques must be carefully reviewed in order to ensure good lifetime under severe load cycling conditions. For nonstationary applications, the reliability of bonded chip contacts is often insufficient. Despite all these limitations, power modules have a promising future. The main reason is that they offer the possibility of further streamlining the construction and assembly in most applications and thus enhancing their economic advantages.

For mass production in the lower-power area, and also in the automobile industry and in consumer goods, intelligent power modules (IPMs) provide the most economical solution. These modules contain intelligent open- and closed-loop control functions, which are tuned for special profiling exercises. The modules additionally contain protection and diagnostic capabilities. Modular internal sensors allow measurements of temperature, currents, and voltages. From this, information status signals for normal operation as well as for fault conditions can be generated.

Cooling

In medium- and low-power inverters forced air ventilation or natural cooling is common. For very large inverters water has shown to be the most effective cooling agent. Water cooling is generally applied for inverters applied in electric traction, and also in electric vehicles. In special cases heat pipe cooling is also suitable for compact construction.

High-Voltage Integrated Circuits

In the low power range, high-voltage integrated circuits (ICs) offer the possibility of drastic reduction in size of inverter design. A monolithic IC contains a power stage (e.g. a three-phase bridge connection of IGBTs and reverse diodes) and also provides control functions such as drive circuits for PWM control at a suitable switching frequency as well as overcur-

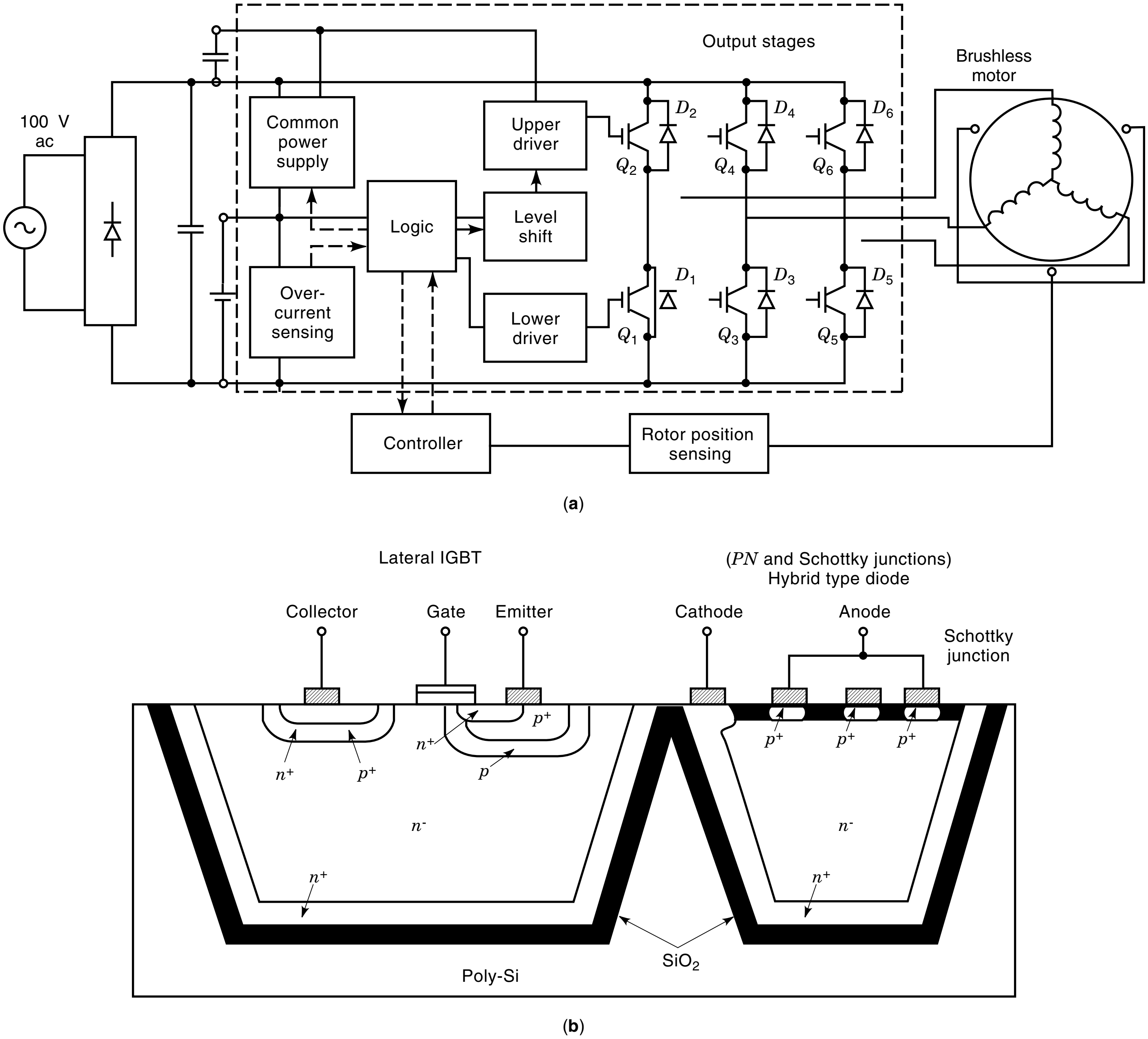


Figure 18. High voltage IC: (a) three-phase monolithic inverter IC (maximum power 50 W, PWM frequency 20 kHz) and (b) cross-section view of IGBT and diode (isolation by a dielectric layer).

rent protection. Figure 18(a) shows the functional block diagram of a high-voltage three-phase monolithic inverter IC for 50 W, with chip size 25 mm^2 (Hitachi). Figure 18(b) shows cross-sectional views of an IGBT and a diode. Isolation of these devices is accomplished by a dielectric layer (25).

Efficiency

The efficiency of inverters depends on the power range. The main losses are caused by the semiconductor valves. The on-state losses are proportional to the forward voltage drop of the semiconductor devices. Unipolar elements have higher forward losses than bipolar elements. The switching losses increase as the applied frequency (carrier or pulse frequency). Additional losses are produced in transformers and reactances. The losses in capacitors are negligible in general.

The efficiency of inverters varies from low to high power in the range of $<90\%$ to $>95\%$. Converter circuits may contain two or more stages of semiconductor circuits (rectifier, inverter, high-frequency link).

APPLICATIONS

Applications of self-commutated inverters cover all areas of electrical engineering. The main application areas are drive technology in industry, traction, and household appliances, and increasingly in office machines and automobiles (26–29). With speed-controlled ac machines such as induction motors and synchronous machines, field-oriented control (vector control) is a very important development in the control of modern inverters. Vector control is now being used in utility and other applications, and integrated with PWM and microprocessor implementation (30). A special modulation method is direct self control (DSC) (31), which allows for flux oriented control and reduction of switching to the minimum required. This control method allows for high dynamics and optimal torque (32). Another important application is in power supplies, from power packs (a few watts) up to installations in the megawatt range.

Other fields are power conditioning, reactive power compensation, and active power filters (33). A special user area is electrotechnology, with middle-frequency inverters for inductive heating, hardening, and melting. See Table 3.

TECHNOLOGICAL TRENDS

The trend is from inverters with thyristors (line- and load-commutated) toward inverters with gate-turnoff semiconductor elements (self-commutated) (34).

Important future progress will be made as a result of the integration of digital information technology, sensor technology, and power electronics. This may well mean customer-oriented, economical solutions and will expand the application field for power electronic solutions considerably.

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