Rajashekara, K., Bhat, A.K.S., Bose, B.K. “Power Electronics”
The Electrical Engineering Handbook
Ed. Richard C. Dorf
Boca Raton: CRC Press LLC, 2000

30

Power Electronics

Kaushik Rajashekara
Delphi Energy & Engine Management Systems

Ashoka K. S. Bhat
University of Victoria

Bimal K. Bose
University of Tennessee

30.1 Power Semiconductor Devices

Kaushik Rajashekara

The modern age of power electronics began with the introduction of thyristors in the late 1950s. Now there are several types of power devices available for high-power and high-frequency applications. The most notable power devices are gate turn-off thyristors, power Darlington transistors, power MOSFETs, and insulated-gate bipolar transistors (IGBTs). Power semiconductor devices are the most important functional elements in all power conversion applications. The power devices are mainly used as switches to convert power from one form to another. They are used in motor control systems, uninterrupted power supplies, high-voltage dc transmission, power supplies, induction heating, and in many other power conversion applications. A review of the basic characteristics of these power devices is presented in this section.

Thyristor and Triac

The thyristor, also called a silicon-controlled rectifier (SCR), is basically a four-layer three-junction *pnpn* device. It has three terminals: anode, cathode, and gate. The device is turned on by applying a short pulse across the gate and cathode. Once the device turns on, the gate loses its control to turn off the device. The turn-off is achieved by applying a **reverse voltage** across the anode and cathode. The thyristor symbol and its volt-ampere characteristics are shown in Fig. 30.1. There are basically two classifications of thyristors: converter grade and inverter grade. The difference between a converter-grade and an inverter-grade thyristor is the low turn-off time (on the order of a few microseconds) for the latter. The converter-grade thyristors are slow type and are used in natural commutation (or phase-controlled) applications. Inverter-grade thyristors are used in forced commutation applications such as dc-dc choppers and dc-ac inverters. The inverter-grade thyristors are turned off by forcing the current to zero using an external commutation circuit. This requires additional commutating components, thus resulting in additional losses in the inverter.

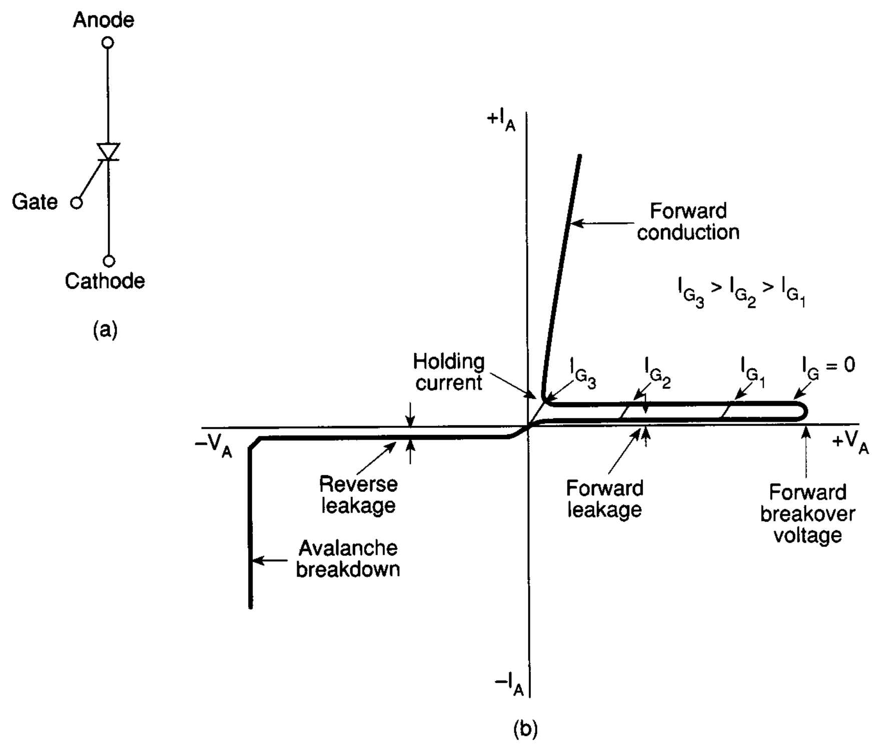


FIGURE 30.1 (a) Thyristor symbol and (b) volt-ampere characteristics. (*Source:* B.K. Bose, *Modern Power Electronics: Evaluation, Technology, and Applications,* p. 5. © 1992 IEEE.)

Thyristors are highly rugged devices in terms of transient currents, ***di/dt,*** and ***dv/dt*** capability. The **forward voltage** drop in thyristors is about 1.5 to 2 V, and even at higher currents of the order of 1000 A, it seldom exceeds 3 V. While the forward voltage determines the on-state power loss of the device at any given current, the switching power loss becomes a dominating factor affecting the device junction temperature at high operating frequencies. Because of this, the maximum switching frequencies possible using thyristors are limited in comparison with other power devices considered in this section.

Thyristors have ***I^2t*** withstand capability and can be protected by fuses. The nonrepetitive surge current capability for thyristors is about 10 times their rated root mean square (rms) current. They must be protected by snubber networks for *dv/dt* and *di/dt* effects. If the specified *dv/dt* is exceeded, thyristors may start conducting without applying a gate pulse. In dc-to-ac conversion applications it is necessary to use an antiparallel diode of similar rating across each main thyristor. Thyristors are available up to 6000 V, 3500 A.

A triac is functionally a pair of converter-grade thyristors connected in antiparallel. The triac symbol and volt-ampere characteristics are shown in Fig. 30.2. Because of the integration, the triac has poor reapplied *dv*/*dt*, poor gate current sensitivity at turn-on, and longer turn-off time. Triacs are mainly used in phase control applications such as in ac regulators for lighting and fan control and in solid-state ac relays.

Gate Turn-Off Thyristor (GTO)

The GTO is a power switching device that can be turned on by a short pulse of gate current and turned off by a reverse gate pulse. This reverse gate current amplitude is dependent on the anode current to be turned off. Hence there is no need for an external commutation circuit to turn it off. Because turn-off is provided by bypassing carriers directly to the gate circuit, its turn-off time is short, thus giving it more capability for high-frequency operation than thyristors. The GTO symbol and turn-off characteristics are shown in Fig. 30.3.

GTOs have the I^2t withstand capability and hence can be protected by semiconductor fuses. For reliable operation of GTOs, the critical aspects are proper design of the gate turn-off circuit and the snubber circuit.

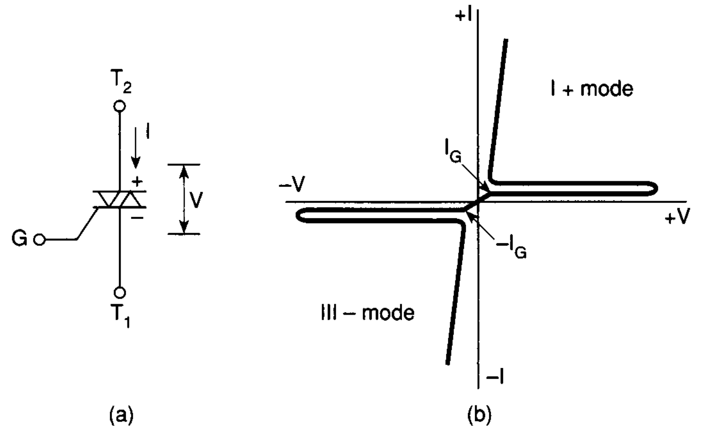


FIGURE 30.2 (a) Triac symbol and (b) volt-ampere characteristics. (*Source:* B.K. Bose, *Modern Power Electronics: Evaluation, Technology, and Applications,* p. 5. © 1992 IEEE.)

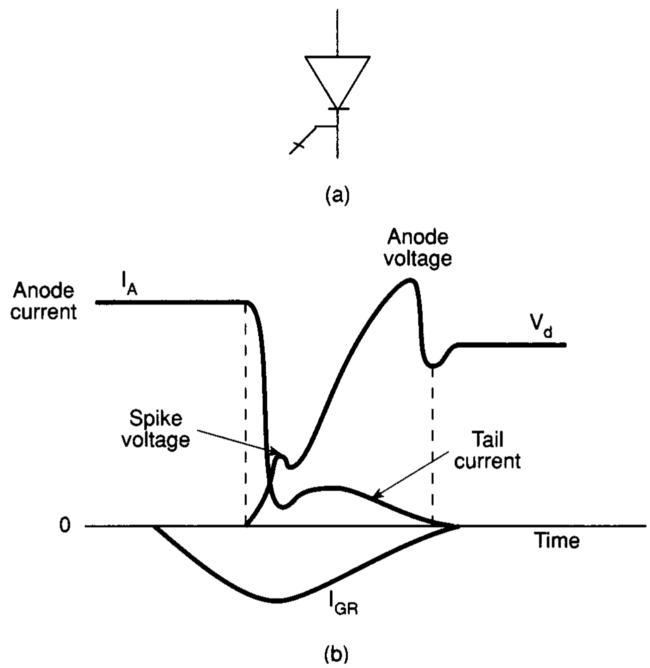


FIGURE 30.3 (a) GTO symbol and (b) turn-off characteristics. (*Source:* B.K. Bose, *Modern Power Electronics: Evaluation, Technology, and Applications,* p. 5. © 1992 IEEE.)

A GTO has a poor turn-off current gain of the order of 4 to 5. For example, a 2000-A peak current GTO may require as high as 500 A of reverse gate current. Also, a GTO has the tendency to latch at temperatures above 125°C. GTOs are available up to about 4500 V, 2500 A.

Reverse-Conducting Thyristor (RCT) and Asymmetrical Silicon-Controlled Rectifier (ASCR)

Normally in inverter applications, a diode in antiparallel is connected to the thyristor for commutation/freewheeling purposes. In RCTs, the diode is integrated with a fast switching thyristor in a single silicon chip. Thus, the number of power devices could be reduced. This integration brings forth a substantial improvement of the static and dynamic characteristics as well as its overall circuit performance.

The RCTs are designed mainly for specific applications such as traction drives. The antiparallel diode limits the reverse voltage across the thyristor to 1 to 2 V. Also, because of the reverse recovery behavior of the diodes, the thyristor may see very high reapplied *dv/dt* when the diode recovers from its reverse voltage. This necessitates use of large RC snubber networks to suppress voltage transients. As the range of application of thyristors and diodes extends into higher frequencies, their reverse recovery charge becomes increasingly important. High reverse recovery charge results in high power dissipation during switching.

The ASCR has a similar forward blocking capability as an inverter-grade thyristor, but it has a limited reverse blocking (about 20–30 V) capability. It has an on-state voltage drop of about 25% less than an inverter-grade thyristor of a similar rating. The ASCR features a fast turn-off time; thus it can work at a higher frequency than an SCR. Since the turn-off time is down by a factor of nearly 2, the size of the commutating components can be halved. Because of this, the switching losses will also be low.

Gate-assisted turn-off techniques are used to even further reduce the turn-off time of an ASCR. The application of a negative voltage to the gate during turn-off helps to evacuate stored charge in the device and aids the recovery mechanisms. This will in effect reduce the turn-off time by a factor of up to 2 over the conventional device.

Power Transistor

Power transistors are used in applications ranging from a few to several hundred kilowatts and switching frequencies up to about 10 kHz. Power transistors used in power conversion applications are generally *npn* type. The power transistor is turned on by supplying sufficient base current, and this base drive has to be maintained throughout its conduction period. It is turned off by removing the base drive and making the base voltage slightly negative (within $-V_{BE(\max)}$). The saturation voltage of the device is normally 0.5 to 2.5 V and increases as the current increases. Hence the on-state losses increase more than proportionately with current. The transistor off-state losses are much lower than the on-state losses because the leakage current of the device is of the order of a few milliamperes. Because of relatively larger switching times, the switching loss significantly increases with switching frequency. Power transistors can block only forward voltages. The reverse peak voltage rating of these devices is as low as 5 to 10 V.

Power transistors do not have I^2t withstand capability. In other words, they can absorb only very little energy before breakdown. Therefore, they cannot be protected by semiconductor fuses, and thus an electronic protection method has to be used.

To eliminate high base current requirements, Darlington configurations are commonly used. They are available in monolithic or in isolated packages. The basic Darlington configuration is shown schematically in Fig. 30.4. The Darlington configuration presents a specific advantage in that it can considerably increase the current switched by the transistor for a given base drive. The $V_{CE(\text{sat})}$ for the Darlington is generally more than that of a single transistor of similar rating with corresponding increase in on-state power loss. During switching, the reverse-biased collector junction may show hot spot breakdown effects that are specified by reverse-bias safe operating area (RBSOA) and forward bias safe operating area (FBSOA). Modern devices with highly interdigited emitter base geometry force more uniform current distribution and therefore considerably improve second breakdown effects. Normally, a well-designed switching aid network constrains the device operation well within the SOAs.

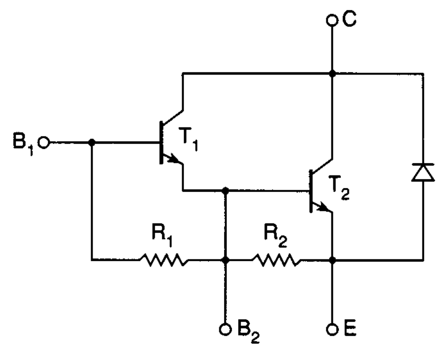


FIGURE 30.4 A two-stage Darlington transistor with bypass diode. (*Source:* B.K. Bose, *Modern Power Electronics: Evaluation, Technology, and Applications,* p. 6. © 1992 IEEE.)

Power MOSFET

Power MOSFETs are marketed by different manufacturers with differences in internal geometry and with different names such as MegaMOS, HEXFET, SIPMOS, and TMOS. They have unique features that make them potentially attractive for switching applications. They are essentially voltage-driven rather than current-driven devices, unlike bipolar transistors.

The gate of a MOSFET is isolated electrically from the source by a layer of silicon oxide. The gate draws only a minute leakage current of the order of nanoamperes. Hence the gate drive circuit is simple and power loss in the gate control circuit is practically negligible. Although in steady state the gate draws virtually no current, this is not so under transient conditions. The gate-to-source and gate-to-drain capacitances have to be charged and discharged appropriately to obtain the desired switching speed, and the drive circuit must have a sufficiently low output impedance to supply the required charging and discharging currents. The circuit symbol of a power MOSFET is shown in Fig. 30.5.

Power MOSFETs are majority carrier devices, and there is no minority carrier storage time. Hence they have exceptionally fast rise and fall times. They are essentially resistive devices when turned on, while bipolar transistors present a more or less constant $V_{CE(\mathrm{sat})}$ over the normal operating range. Power dissipation in MOSFETs is $Id^2R_{DS(\mathrm{on})}$, and in bipolars it is $I_C V_{CE(\mathrm{sat})}$. At low currents, therefore, a power MOSFET may have a lower conduction loss than a comparable bipolar device, but at higher currents, the conduction loss will exceed that of bipolars. Also, the $R_{DS(\mathrm{on})}$ increases with temperature.

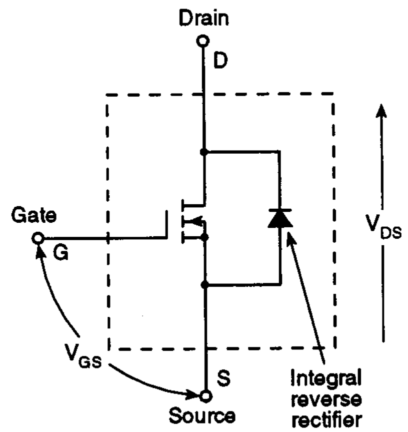


FIGURE 30.5 Power MOSFET circuit symbol. (*Source:* B.K. Bose, *Modern Power Electronics: Evaluation, Technology, and Applications,* p. 7. © 1992 IEEE.)

An important feature of a power MOSFET is the absence of a secondary breakdown effect, which is present in a bipolar transistor, and as a result, it has an extremely rugged switching performance. In MOSFETs, $R_{DS(\mathrm{on})}$ increases with temperature, and thus the current is automatically diverted away from the hot spot. The drain body junction appears as an antiparallel diode between source and drain. Thus power MOSFETs will not support voltage in the reverse direction. Although this inverse diode is relatively fast, it is slow by comparison with the MOSFET. Recent devices have the diode recovery time as low as 100 ns. Since MOSFETs cannot be protected by fuses, an electronic protection technique has to be used.

With the advancement in MOS technology, ruggedized MOSFETs are replacing the conventional MOSFETs. The need to ruggedize power MOSFETs is related to device reliability. If a MOSFET is operating within its specification range at all times, its chances for failing catastrophically are minimal. However, if its absolute maximum rating is exceeded, failure probability increases dramatically. Under actual operating conditions, a MOSFET may be subjected to transients — either externally from the power bus supplying the circuit or from the circuit itself due, for example, to inductive kicks going beyond the absolute maximum ratings. Such conditions are likely in almost every application, and in most cases are beyond a designer's control. Rugged devices are made to be more tolerant for over-voltage transients. Ruggedness is the ability of a MOSFET to operate in an environment of dynamic electrical stresses, without activating any of the parasitic bipolar junction transistors. The rugged device can withstand higher levels of diode recovery *dv/dt* and static *dv/dt.*

Insulated-Gate Bipolar Transistor (IGBT)

The IGBT has the high input impedance and high-speed characteristics of a MOSFET with the conductivity characteristic (low saturation voltage) of a bipolar transistor. The IGBT is turned on by applying a positive voltage between the gate and emitter and, as in the MOSFET, it is turned off by making the gate signal zero or slightly negative. The IGBT has a much lower voltage drop than a MOSFET of similar ratings. The structure of an IGBT is more like a thyristor and MOSFET. For a given IGBT, there is a critical value of collector current that will cause a large enough voltage drop to activate the thyristor. Hence, the device manufacturer specifies the peak allowable collector current that can flow without latch-up occurring. There is also a corresponding gate source voltage that permits this current to flow that should not be exceeded.

Like the power MOSFET, the IGBT does not exhibit the secondary breakdown phenomenon common to bipolar transistors. However, care should be taken not to exceed the maximum power dissipation and specified maximum junction temperature of the device under all conditions for guaranteed reliable operation. The on-state voltage of the IGBT is heavily dependent on the gate voltage. To obtain a low on-state voltage, a sufficiently high gate voltage must be applied.

In general, IGBTs can be classified as punch-through (PT) and nonpunch-through (NPT) structures, as shown in Fig. 30.6. In the PT IGBT, an N^+ buffer layer is normally introduced between the P^+ substrate and the N^- epitaxial layer, so that the whole N^- drift region is depleted when the device is blocking the off-state voltage, and the electrical field shape inside the N^- drift region is close to a rectangular shape. Because a shorter N^- region can be used in the punch-through IGBT, a better trade-off between the forward voltage drop and turn-off time can be achieved. PT IGBTs are available up to about 1200 V.

High voltage IGBTs are realized through non-punch-through process. The devices are built on a N^- wafer substrate which serves as the N^- base drift region. Experimental NPT IGBTs of up to about 4 KV have been reported in the literature. NPT IGBTs are more robust than PT IGBTs particularly under short circuit conditions. But NPT IGBTs have a higher forward voltage drop than the PT IGBTs.

The PT IGBTs cannot be as easily paralleled as MOSFETs. The factors that inhibit current sharing of parallel-connected IGBTs are (1) on-state current unbalance, caused by V_{CE}(sat) distribution and main circuit wiring resistance distribution, and (2) current unbalance at turn-on and turn-off, caused by the switching time difference of the parallel connected devices and circuit wiring inductance distribution. The NPT IGBTs can be paralleled because of their positive temperature coefficient property.

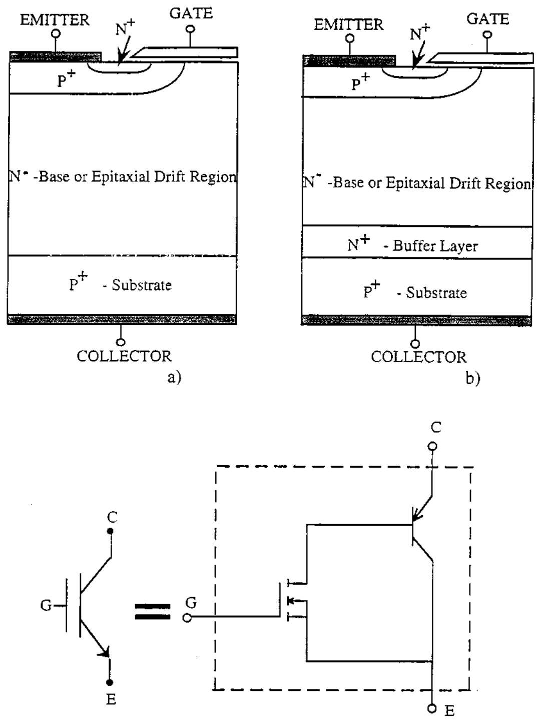


FIGURE 30.6 Nonpunch-through IGBT, (b) Punch-through IGBT, (c) IGBT equivalent circuit.

MOS-Controlled Thyristor (MCT)

The MCT is a new type of power semiconductor device that combines the capabilities of thyristor voltage and current with MOS gated turn-on and turn-off. It is a high power, high frequency, low conduction drop and a rugged device, which is more likely to be used in the future for medium and high power applications. A cross sectional structure of a p-type MCT with its circuit schematic is shown in Fig. 30.7. The MCT has a thyristor type structure with three junctions and PNPN layers between the anode and cathode. In a practical MCT, about 100,000 cells similar to the one shown are paralleled to achieve the desired current rating. MCT is turned on by a negative voltage pulse at the gate with respect to the anode, and is turned off by a positive voltage pulse.

The MCT was announced by the General Electric R & D Center on November 30, 1988. Harris Semiconductor Corporation has developed two generations of p-MCTs. Gen-1 p-MCTs are available at 65 A/1000 V and 75A/600 V with peak controllable current of 120 A. Gen-2 p-MCTs are being developed at similar current and voltage ratings, with much improved turn-on capability and switching speed. The reason for developing p-MCT is the fact that the current density that can be turned off is 2 or 3 times higher than that of an n-MCT; but n-MCTs are the ones needed for many practical applications. Harris Semiconductor Corporation is in the process of developing n-MCTs, which are expected to be commercially available during the next one to two years.

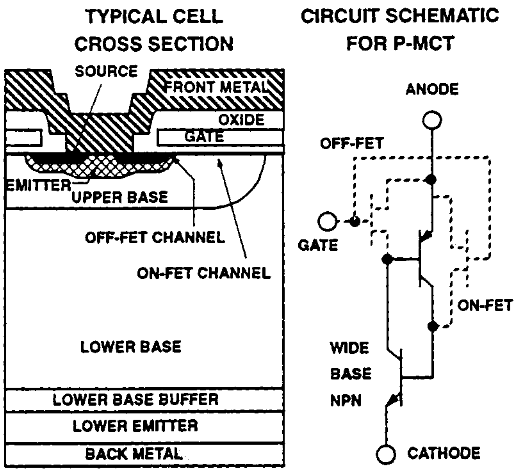


FIGURE 30.7 (*Source:* Harris Semiconductor, *User's Guide of MOS Controlled Thyristor,* With permission.)

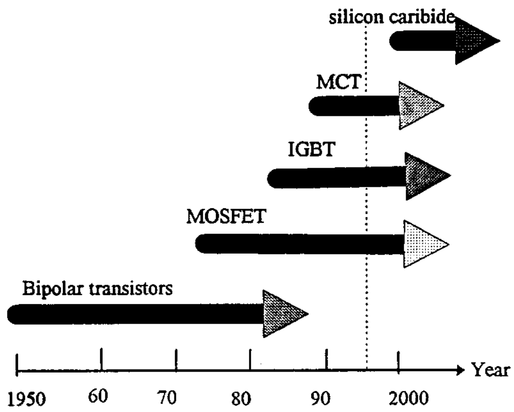


FIGURE 30.8 Current and future pwer semiconductor devices development direction (*Source:* A.Q. Huang, *Recent Developments of Power Semiconductor Devices,* VPEC Seminar Proceedings, pp. 1–9. With permission.)

The advantage of an MCT over-IGBT is its low forward voltage drop. N-type MCTs will be expected to have a similar forward voltage drop, but with an improved reverse bias safe operating area and switching speed. MCTs have relatively low switching times and storage time. The MCT is capable of high current densities and blocking voltages in both directions. Since the power gain of an MCT is extremely high, it could be driven directly from logic gates. An MCT has high *di/dt* (of the order of 2500 A/μs) and high *dv/dt* (of the order of 20,000 V/μs) capability.

The MCT, because of its superior characteristics, shows a tremendous possibility for applications such as motor drives, uninterrupted power supplies, static VAR compensators, and high power active power line conditioners.

The current and future power semiconductor devices developmental direction is shown in Fig. 30.8. High temperature operation capability and low forward voltage drop operation can be obtained if silicon is replaced by silicon carbide material for producing power devices. The silicon carbide has a higher band gap than silicon. Hence higher breakdown voltage devices could be developed. Silicon carbide devices have excellent switching characteristics and stable blocking voltages at higher temperatures. But the silicon carbide devices are still in the very early stages of development.

Defining Terms

di/dt limit: Maximum allowed rate of change of current through a device. If this limit is exceeded, the device may not be guaranteed to work reliably.

dv/dt: Rate of change of voltage withstand capability without spurious turn-on of the device.

Forward voltage: The voltage across the device when the anode is positive with respect to the cathode.

I^2t: Represents available thermal energy resulting from current flow.

Reverse voltage: The voltage across the device when the anode is negative with respect to the cathode.

Related Topic

5.1 Diodes and Rectifiers

References

B.K. Bose, *Modern Power Electronics: Evaluation, Technology, and Applications,* New York: IEEE Press, 1992.

Harris Semiconductor, *User's Guide of MOS Controlled Thyristor.*

A.Q. Huang, *Recent Developments of Power Semiconductor Devices,* VPEC Seminar Proceedings, pp. 1–9, September 1995.

N. Mohan and T. Undeland, *Power Electronics: Converters, Applications, and Design,* New York: John Wiley & Sons, 1995.

J. Wojslawowicz, "Ruggedized transistors emerging as power MOSFET standard-bearers," *Power Technics Magazine,* pp. 29–32, January 1988.

Further Information

B.M. Bird and K.G. King, *An Introduction to Power Electronics,* New York: Wiley-Interscience, 1984.

R. Sittig and P. Roggwiller, *Semiconductor Devices for Power Conditioning,* New York: Plenum, 1982.

V.A.K. Temple, "Advances in MOS controlled thyristor technology and capability," *Power Conversion*, pp. 544–554, Oct. 1989.

B.W. Williams, *Power Electronics, Devices, Drivers and Applications,* New York: John Wiley, 1987.

30.2 Power Conversion

Kaushik Rajashekara

Power conversion deals with the process of converting electric power from one form to another. The power electronic apparatuses performing the power conversion are called *power converters.* Because they contain no moving parts, they are often referred to as *static* power converters. The power conversion is achieved using power semiconductor devices, which are used as switches. The power devices used are SCRs (silicon controlled rectifiers, or thyristors), triacs, power transistors, power MOSFETs, insulated gate bipolar transistors (IGBTs), and MCTs (MOS-controlled thyristors). The power converters are generally classified as:

1. ac-dc converters (phase-controlled converters)
2. direct ac-ac converters (cycloconverters)
3. dc-ac converters (inverters)
4. dc-dc converters (choppers, buck and boost converters)

AC-DC Converters

The basic function of a **phase-controlled converter** is to convert an alternating voltage of variable amplitude and frequency to a variable dc voltage. The power devices used for this application are generally **SCR**s. The average value of the output voltage is controlled by varying the conduction time of the SCRs. The turn-on of the SCR is achieved by providing a gate pulse when it is forward-biased. The turn-off is achieved by the **commutation** of current from one device to another at the instant the incoming ac voltage has a higher instantaneous potential than that of the outgoing wave. Thus there is a natural tendency for current to be commutated from the outgoing to the incoming SCR, without the aid of any external commutation circuitry. This commutation process is often referred to as *natural commutation.*

A single-phase half-wave converter is shown in Fig. 30.9. When the SCR is turned on at an angle α, full supply voltage (neglecting the SCR drop) is applied to the load. For a purely resistive load, during the positive half cycle, the output voltage waveform follows the input ac voltage waveform. During the negative half cycle, the SCR is turned off. In the case of inductive load, the energy stored in the inductance causes the current to flow in the load circuit even after the reversal of the supply voltage, as shown in Fig. 30.9(b). If there is no freewheeling diode D_F, the load current is discontinuous. A freewheeling diode is connected across the load to turn off the SCR as soon as the input voltage polarity reverses, as shown in Fig. 30.9(c). When the SCR is off, the load current will freewheel through the diode. The power flows from the input to the load only when the SCR is conducting. If there is no freewheeling diode, during the negative portion of the supply voltage, SCR returns the energy stored in the load inductance to the supply. The freewheeling diode improves the input power factor.

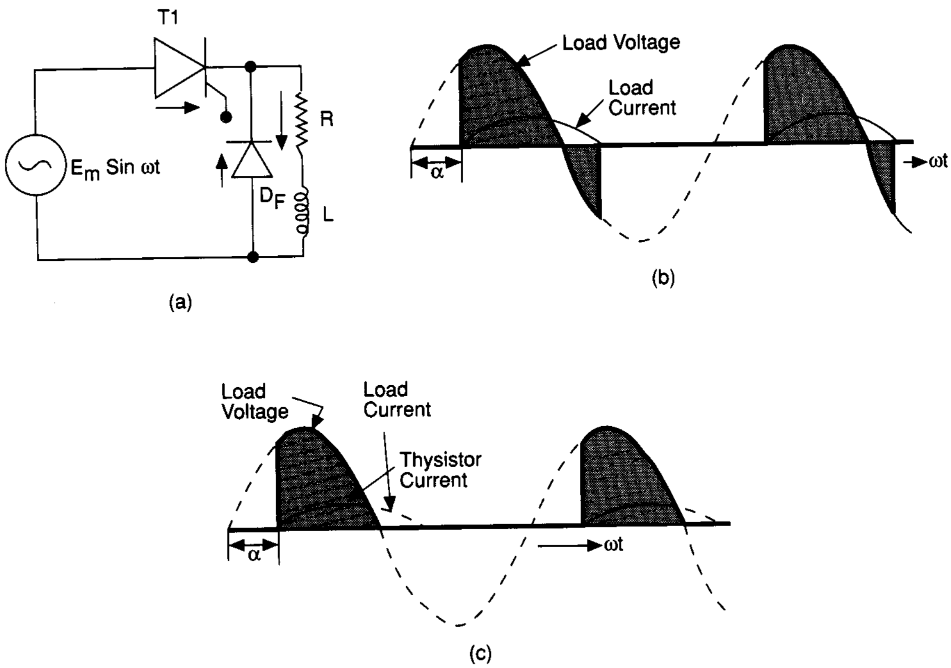


FIGURE 30.9 Single-phase half-wave converter with freewheeling diode. (a) Circuit diagram; (b) waveform for inductive load with no freewheeling diode; (c) waveform with freewheeling diode.

The controlled full-wave dc output may be obtained by using either a center tap transformer (Fig. 30.10) or by bridge configuration (Fig. 30.11). The bridge configuration is often used when a transformer is undesirable and the magnitude of the supply voltage properly meets the load voltage requirements. The average output voltage of a single-phase full-wave converter for continuous current conduction is given by

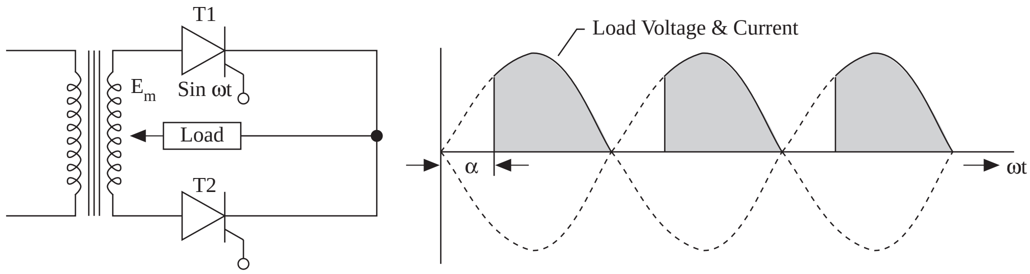
$$v_{d\alpha} = 2\frac{E_m}{\pi}\cos\alpha$$

where E_m is the peak value of the input voltage and α is the firing angle. The output voltage of a single-phase bridge circuit is the same as that shown in Fig. 30.10. Various configurations of the single-phase bridge circuit can be obtained if, instead of four SCRs, two diodes and two SCRs are used, with or without freewheeling diodes.

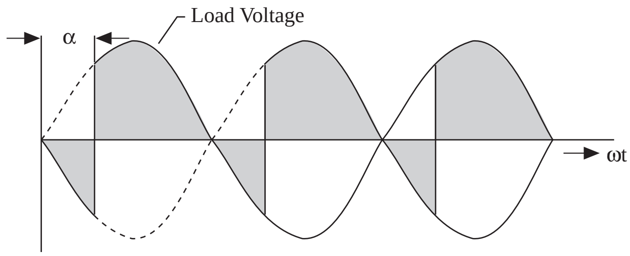
A three-phase full-wave converter consisting of six thyristor switches is shown in Fig. 30.12(a). This is the most commonly used three-phase bridge configuration. Thyristors T_1, T_3, and T_5 are turned on during the positive half cycle of the voltages of the phases to which they are connected, and thyristors T_2, T_4, and T_6 are turned on during the negative half cycle of the phase voltages. The reference for the angle in each cycle is at the crossing points of the phase voltages. The ideal output voltage, output current, and input current waveforms are shown in Fig. 30.12(b). The output dc voltage is controlled by varying the firing angle α. The average output voltage under continuous current conduction operation is given by

$$v_o = \frac{3\sqrt{3}}{\pi}E_m\cos\alpha$$

where E_m is the peak value of the phase voltage. At $\alpha = 90°$, the output voltage is zero. For $0 < \alpha < 90°$, v_o is positive and power flows from ac supply to the load. For $90° < \alpha < 180°$, v_o is negative and the converter operates in the inversion mode. If the load is a dc motor, the power can be transferred from the motor to the ac supply, a process known as *regeneration.*



(a) For Resistive Load



(b) For Resistive-Inductive Load (with continuous current conduction)

FIGURE 30.10 Single-phase full-wave converter with transformer.

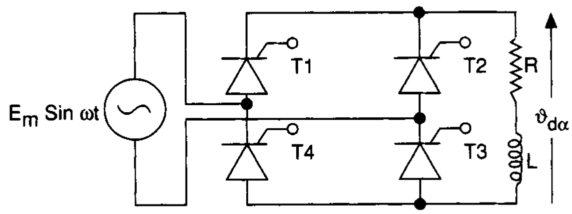


FIGURE 30.11 Single-phase bridge converter.

In Fig. 30.12(a), the top or bottom thyristors could be replaced by diodes. The resulting topology is called a *thyristor semiconverter*. With this configuration, the input power factor is improved, but the regeneration is not possible.

Cycloconverters

Cycloconverters are direct ac-to-ac frequency changers. The term *direct conversion* means that the energy does not appear in any form other than the ac input or ac output. The output frequency is lower than the input frequency and is generally an integral multiple of the input frequency. A cycloconverter permits energy to be fed back into the utility network without any additional measures. Also, the phase sequence of the output voltage can be easily reversed by the control system. Cycloconverters have found applications in aircraft systems and industrial drives. These cycloconverters are suitable for synchronous and induction motor control. The operation of the cycloconverter is illustrated in Section 30.4 of this chapter.

DC-to-AC Converters

The dc-to-ac converters are generally called *inverters*. The ac supply is first converted to dc, which is then converted to a variable-voltage and variable-frequency power supply. This generally consists of a three-phase bridge connected to the ac power source, a dc link with a filter, and the three-phase inverter bridge connected

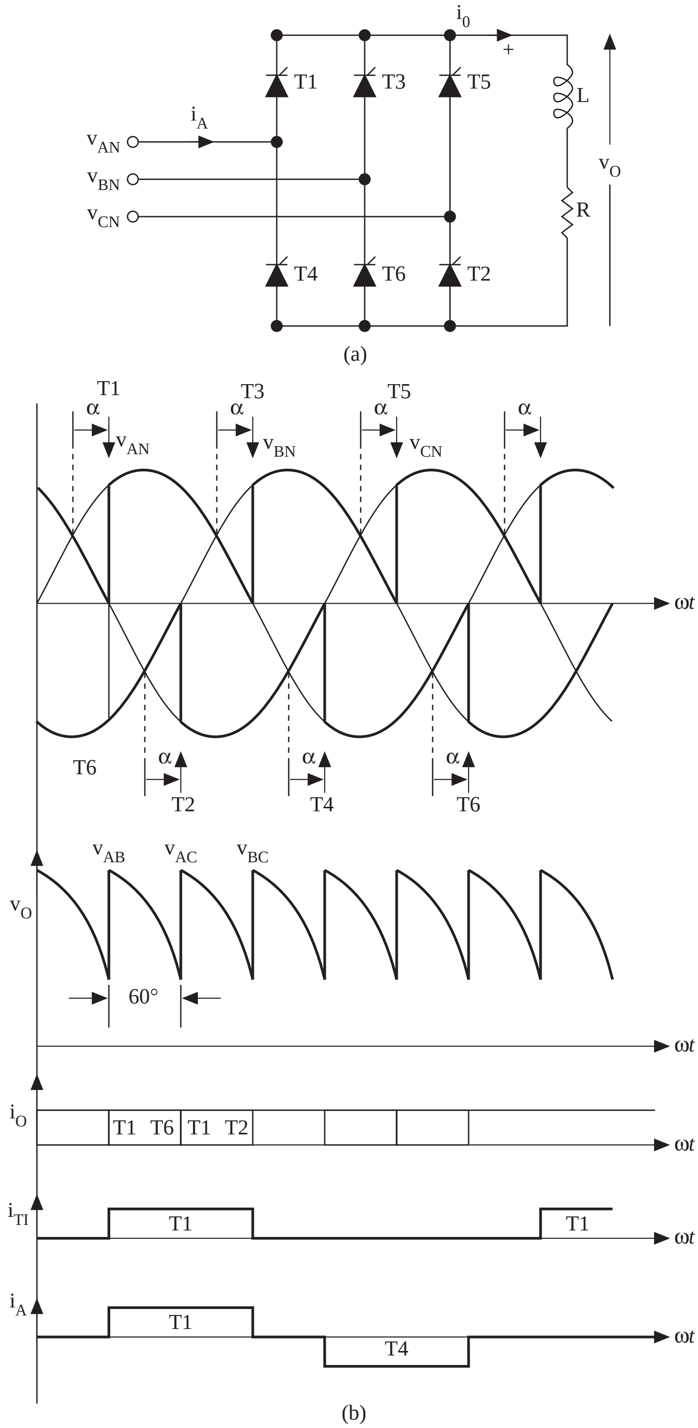


FIGURE 30.12 (a) Three-phase thyristor full bridge configuration; (b) output voltage and current waveforms.

to the load. In the case of battery-operated systems, there is no intermediate dc link. Inverters can be classified as voltage source inverters (VSIs) and current source inverters (CSIs). A voltage source inverter is fed by a stiff dc voltage, whereas a current source inverter is fed by a stiff current source. A voltage source can be converted to a current source by connecting a series inductance and then varying the voltage to obtain the desired current.

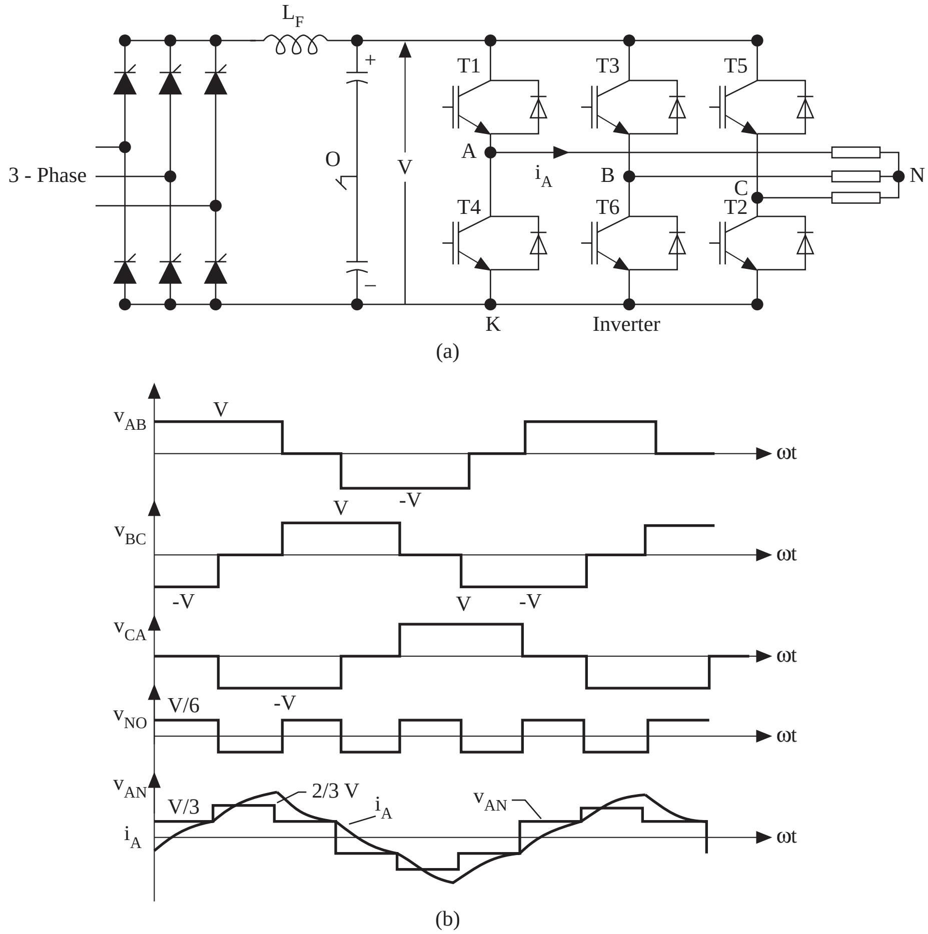


FIGURE 30.13 (a) Three-phase converter and voltage source inverter configuration; (b) three-phase square-wave inverter waveforms.

A VSI can also be operated in current-controlled mode, and similarly a CSI can also be operated in the voltage-control mode. The inverters are used in variable frequency ac motor drives, uninterrupted power supplies, induction heating, static VAR compensators, etc.

Voltage Source Inverter

A three-phase voltage source inverter configuration is shown in Fig. 30.13(a). The VSIs are controlled either in square-wave mode or in pulsewidth-modulated (PWM) mode. In square-wave mode, the frequency of the output voltage is controlled within the inverter, the devices being used to switch the output circuit between the plus and minus bus. Each device conducts for 180 degrees, and each of the outputs is displaced 120 degrees to generate a six-step waveform, as shown in Fig. 30.13(b). The amplitude of the output voltage is controlled by varying the dc link voltage. This is done by varying the firing angle of the thyristors of the three-phase bridge converter at the input. The square-wave-type VSI is not suitable if the dc source is a battery. The six-step output voltage is rich in harmonics and thus needs heavy filtering.

In PWM inverters, the output voltage and frequency are controlled within the inverter by varying the width of the output pulses. Hence at the front end, instead of a phase-controlled thyristor converter, a diode bridge rectifier can be used. A very popular method of controlling the voltage and frequency is by sinusoidal pulsewidth modulation. In this method, a high-frequency triangle carrier wave is compared with a three-phase sinusoidal waveform, as shown in Fig. 30.14. The power devices in each phase are switched on at the intersection of sine

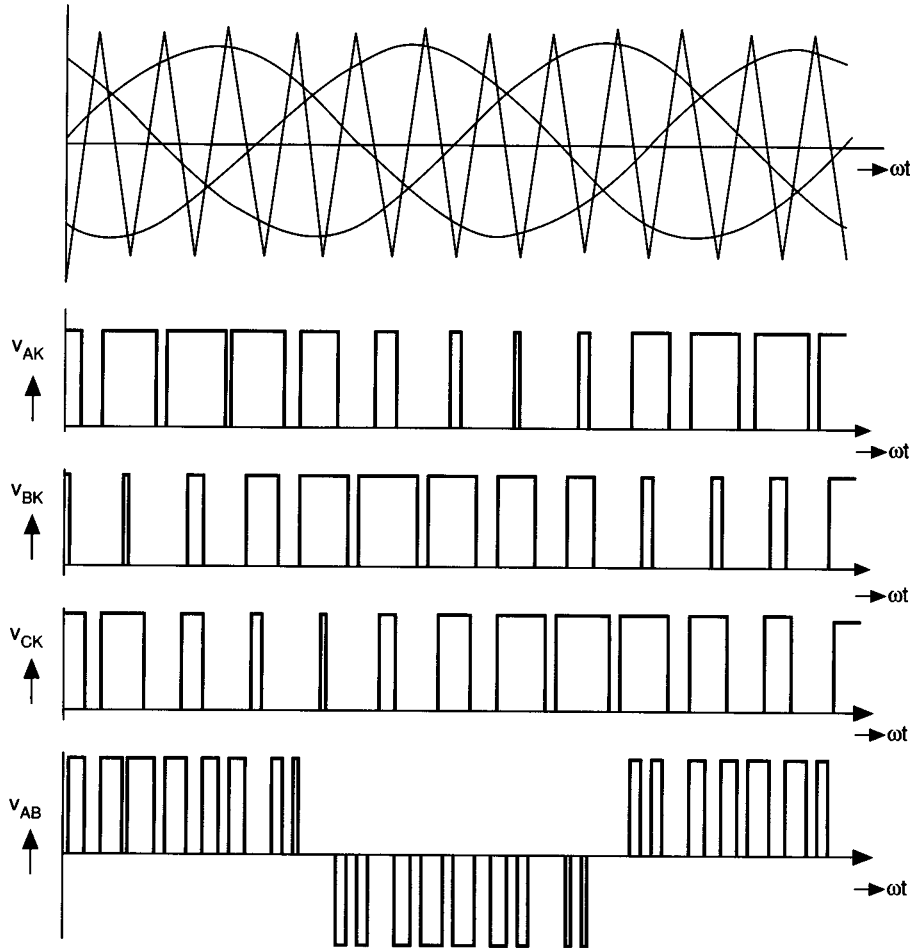


FIGURE 30.14 Three-phase sinusoidal PWM inverter waveforms.

and triangle waves. The amplitude and frequency of the output voltage are varied, respectively, by varying the amplitude and frequency of the reference sine waves. The ratio of the amplitude of the sine wave to the amplitude of the carrier wave is called the *modulation index.*

The harmonic components in a PWM wave are easily filtered because they are shifted to a higher-frequency region. It is desirable to have a high ratio of carrier frequency to fundamental frequency to reduce the harmonics of lower-frequency components. There are several other PWM techniques mentioned in the literature. The most notable ones are selected harmonic elimination, hysteresis controller, and space vector PWM technique.

In inverters, if SCRs are used as power switching devices, an external forced commutation circuit has to be used to turn off the devices. Now, with the availability of **IGBT**s above 1000-A, 1000-V ratings, they are being used in applications up to 300-kW motor drives. Above this power rating, GTOs are generally used. Power Darlington transistors, which are available up to 800 A, 1200 V, could also be used for inverter applications.

Current Source Inverter

Contrary to the voltage source inverter where the voltage of the dc link is imposed on the motor windings, in the current source inverter the current is imposed into the motor. Here the amplitude and phase angle of the motor voltage depend on the load conditions of the motor. The current source inverter is described in detail in Section 30.4.

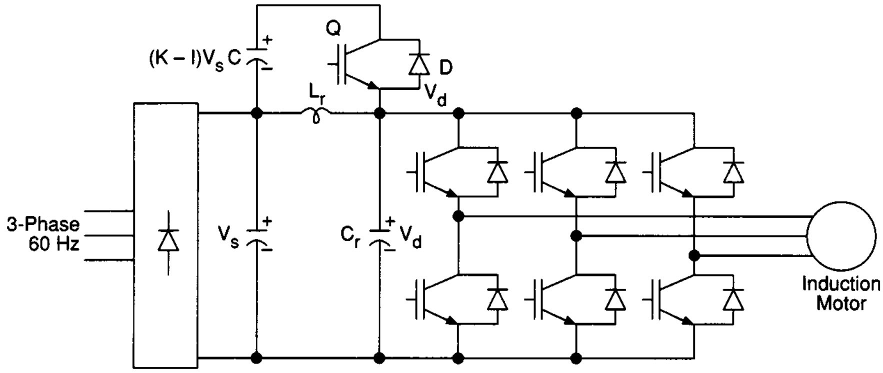


FIGURE 30.15 Resonant dc-link inverter system with active voltage clamping.

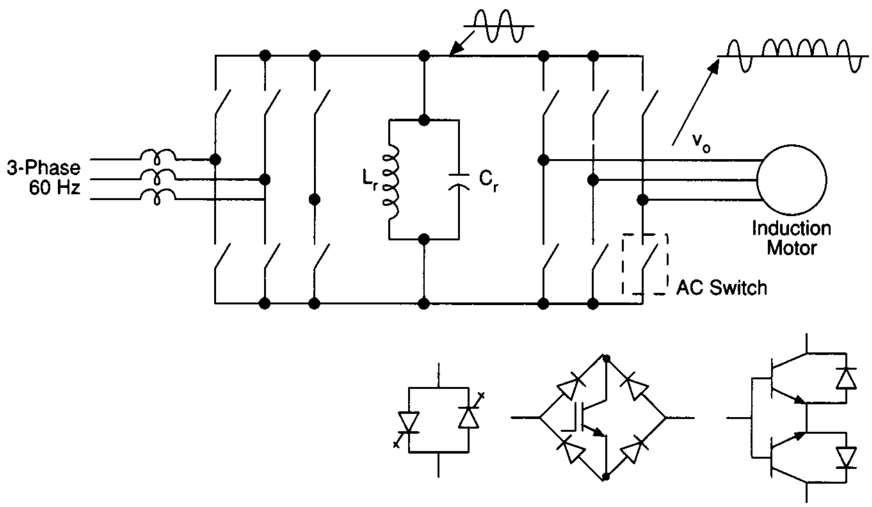


FIGURE 30.16 Resonant ac-link converter system showing configuration of ac switches.

Resonant-Link Inverters

The use of resonant switching techniques can be applied to inverter topologies to reduce the switching losses in the power devices. They also permit high switching frequency operation to reduce the size of the magnetic components in the inverter unit. In the resonant dc-link inverter shown in Fig. 30.15, a resonant circuit is added at the inverter input to convert a fixed dc voltage to a pulsating dc voltage. This resonant circuit enables the devices to be turned on and turned off during the zero voltage interval. Zero voltage or zero current switching is often termed *soft switching*. Under soft switching, the switching losses in the power devices are almost eliminated. The electromagnetic interference (EMI) problem is less severe because resonant voltage pulses have lower *dv/dt* compared to those of hard-switched PWM inverters. Also, the machine insulation is less stretched because of lower *dv/dt* resonant voltage pulses. In Fig. 30.15, all the inverter devices are turned on simultaneously to initiate a resonant cycle. The commutation from one device to another is initiated at the zero dc-link voltage. The inverter output voltage is formed by the integral numbers of quasi-sinusoidal pulses. The circuit consisting of devices *Q*, *D*, and the capacitor *C* acts as an active clamp to limit the dc voltage to about 1.4 times the diode rectifier voltage V_s.

There are several other topologies of resonant link inverters mentioned in the literature. There are also resonant link ac-ac converters based on bidirectional ac switches, as shown in Fig. 30.16. These resonant link converters find applications in ac machine control and uninterrupted power supplies, induction heating, etc. The resonant link inverter technology is still in the development stage for industrial applications.

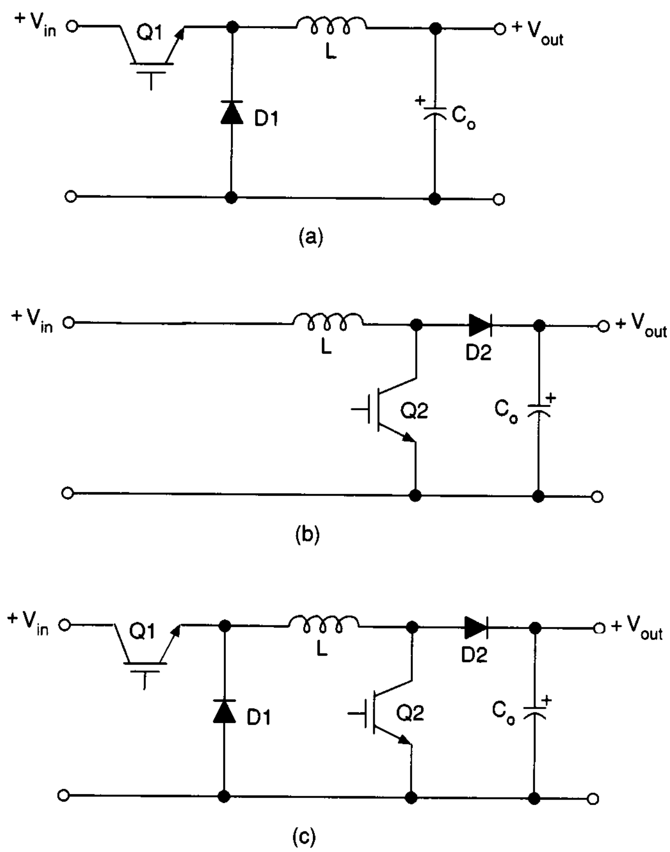


FIGURE 30.17 DC-DC converter configurations: (a) buck converter; (b) boost converter; (c) buck-boost converter.

DC-DC Converters

DC-dc converters are used to convert unregulated dc voltage to regulated or variable dc voltage at the output. They are widely used in switch-mode dc power supplies and in dc motor drive applications. In dc motor control applications, they are called *chopper-controlled drives*. The input voltage source is usually a battery or derived from an ac power supply using a diode bridge rectifier. These converters are generally either hard-switched PWM types or soft-switched resonant-link types. There are several dc-dc converter topologies, the most common ones being buck converter, boost converter, and buck-boost converter, shown in Fig. 30.17.

Buck Converter

A buck converter is also called a *step-down* converter. Its principle of operation is illustrated by referring to Fig. 30.17(a). The IGBT acts as a high-frequency switch. The IGBT is repetitively closed for a time t_{on} and opened for a time t_{off}. During t_{on}, the supply terminals are connected to the load, and power flows from supply to the load. During t_{off}, load current flows through the freewheeling diode D_1, and the load voltage is ideally zero. The average output voltage is given by

$$V_{out} = DV_{in}$$

where D is the **duty cycle** of the switch and is given by $D = t_{on}/T$, where T is the time for one period. $1/T$ is the switching frequency of the power device IGBT.

Boost Converter

A boost converter is also called a *step-up* converter. Its principle of operation is illustrated by referring to Fig. 30.17(b). This converter is used to produce higher voltage at the load than the supply voltage. When the

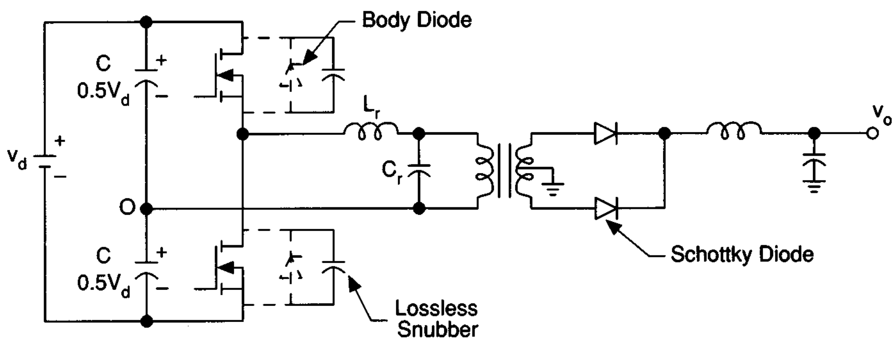


FIGURE 30.18 Resonant-link dc-dc converter.

power switch is on, the inductor is connected to the dc source and the energy from the supply is stored in it. When the device is off, the inductor current is forced to flow through the diode and the load. The induced voltage across the inductor is negative. The inductor adds to the source voltage to force the inductor current into the load. The output voltage is given by

$$V_{out} = \frac{V_{in}}{1 - D}$$

Thus for variation of D in the range $0 < D < 1$, the load voltage V_{out} will vary in the range $V_{in} < V_{out} < \infty$.

Buck-Boost Converter

A buck-boost converter can be obtained by the cascade connection of the buck and the boost converter. The steady-state output voltage V_{out} is given by

$$V_{out} = V_{in} \frac{D}{1 - D}$$

This allows the output voltage to be higher or lower than the input voltage, based on the duty cycle D. A typical buck-boost converter topology is shown in Fig. 30.17(c). When the power device is turned on, the input provides energy to the inductor and the diode is reverse biased. When the device is turned off, the energy stored in the inductor is transferred to the output. No energy is supplied by the input during this interval. In dc power supplies, the output capacitor is assumed to be very large, which results in a constant output voltage. In dc drive systems, the chopper is operated in step-down mode during motoring and in step-up mode during regeneration operation.

Resonant-Link DC-DC Converters

The use of resonant converter topologies would help to reduce the switching losses in dc-dc converters and enable the operation at switching frequencies in the megahertz range. By operating at high frequencies, the size of the power supplies could be reduced. There are several types of resonant converter topologies. The most popular configuration is shown in Fig. 30.18. The dc power is converted to high-frequency alternating power using the MOSFET half-bridge inverter. The resonant capacitor voltage is transformer-coupled, rectified using the two Schottky diodes, and then filtered to get output dc voltage. The output voltage is regulated by control of the inverter switching frequency.

Instead of parallel loading as in Fig. 30.18, the resonant circuit can be series-loaded; that is, the transformer in the output circuit can be placed in series with the tuned circuit. The series resonant circuit provides the short-circuit limiting feature.

There are other forms of resonant converter topologies mentioned in the literature such as quasi-resonant converters and multiresonant converters. These resonant converter topologies find applications in high-density power supplies.

Defining Terms

Commutation: Process of transferring the current from one power device to another.
Duty cycle: Ratio of the on-time of a switch to the switching period.
Full-wave control: Both the positive and negative half cycle of the waveforms are controlled.
IGBT: Insulated-gate bipolar transistor.
Phase-controlled converter: Converter in which the power devices are turned off at the natural crossing of zero voltage in ac to dc conversion applications.
SCR: Silicon-controlled rectifier.

Related Topics

33.2 Heat Transfer Fundamentals • 61.3 High-Voltage Direct-Current Transmission

References

B.K. Bose, *Modern Power Electronics,* New York: IEEE Press, 1992.
Motorola, *Linear/Switchmode Voltage Regulator Handbook,* 1989.
K.S. Rajashekara, H. Le-Huy, et al., "Resonant DC Link Inverter-Fed AC Machines Control," IEEE Power Electronics Specialists Conference, 1987, pp. 491–496.
P.C. Sen, *Thyristor DC Drives,* New York: John Wiley, 1981.
G. Venkataramanan and D. Divan, "Pulse Width Modulation with Resonant DC Link Converters," IEEE IAS Annual Meeting, 1990, pp. 984–990.

Further Information

B.K. Bose, *Power Electronics & AC Drives,* Englewood Cliffs, N.J.: Prentice-Hall, 1986.
R. Hoft, *Semiconductor Power Electronics,* New York: Van Nostrand Reinhold, 1986.
B.R. Pelly, *Thyristor Phase Controlled Converters and Cycloconverters,* New York: Wiley-Interscience, 1971.
A.I. Pressman, *Switching and Linear Power Supply, Power Converter Design,* Carmel, Ind.: Hayden Book Company, 1977.
M.H. Rashid, *Power Electronics, Circuits, Devices and Applications,* Englewood Cliffs, N.J.: Prentice-Hall, 1988.

30.3 Power Supplies

Ashoka K. S. Bhat

Power supplies are used in many industrial and aerospace applications and also in consumer products. Some of the requirements of power supplies are small size, light weight, low cost, and high power conversion efficiency. In addition to these, some power supplies require the following: electrical isolation between the source and load, low harmonic distortion for the input and output waveforms, and high power factor (PF) if the source is ac voltage. Some special power supplies require controlled direction of power flow.

Basically two types of power supplies are required: dc power supplies and ac power supplies. The output of dc power supplies is regulated or controllable dc, whereas the output for ac power supplies is ac. The input to these power supplies can be ac or dc.

DC Power Supplies

If an ac source is used, then ac-to-dc **converters** explained in Section 30.2 can be used. In these converters, electrical isolation can only be provided by bulky line frequency transformers. The ac source can be rectified with a diode rectifier to get an uncontrolled dc, and then a dc-to-dc converter can be used to get a controlled dc output. Electrical isolation between the input source and the output load can be provided in the dc-to-dc converter using a high-frequency (HF) transformer. Such HF transformers have small size, light weight, and low cost compared to bulky line frequency transformers. Whether the input source is dc (e.g., battery) or ac, dc-to-dc converters form an important part of dc power supplies, and they are explained in this subsection.

DC power supplies can be broadly classified as linear and switching power supplies.

A *linear power supply* is the oldest and simplest type of power supply. The output voltage is **regulated** by dropping the extra input voltage across a series transistor (therefore, also referred to as a series regulator). They have very small output ripple, theoretically zero noise, large hold-up time (typically 1–2 ms), and fast response. Linear power supplies have the following disadvantages: very low efficiency, electrical isolation can only be on 60-Hz ac side, larger volume and weight, and, in general, only a single output possible. However, they are still used in very small regulated power supplies and in some special applications (e.g., magnet power supplies). Three terminal linear regulator integrated circuits (ICs) are readily available (e.g., µA7815 has +15-V, 1-A output), are easy to use, and have built-in load short-circuit protection.

Switching power supplies use power semiconductor switches in the *on* and *off* switching states resulting in high efficiency, small size, and light weight. With the availability of fast switching devices, HF magnetics and capacitors, and high-speed control ICs, switching power supplies have become very popular. They can be further classified as **pulsewidth-modulated (PWM) converters** and **resonant converters**, and they are explained below.

Pulsewidth-Modulated Converters

These converters employ square-wave pulsewidth modulation to achieve voltage regulation. The average output voltage is varied by varying the duty cycle of the power semiconductor switch. The voltage waveform across the switch and at the output are square wave in nature [refer to Fig. 30.13(b)] and they generally result in higher switching losses when the switching frequency is increased. Also, the switching stresses are high with the generation of large electromagnetic interference (EMI), which is difficult to filter. However, these converters are easy to control, well understood, and have wide load control range.

The methods of control of PWM converters are discussed next.

The Methods of Control. The PWM converters operate with a fixed-frequency, variable duty cycle. Depending on the duty cycle, they can operate in either continuous current mode (CCM) or discontinuous current mode (DCM). If the current through the output inductor never reaches zero (refer to Fig. 30.13), then the converter operates in CCM; otherwise DCM occurs.

The three possible control methods [Severns and Bloom, 1988; Hnatek, 1981; Unitrode Corporation, 1984; Motorola, 1989; Philips Semiconductors, 1991] are briefly explained below.

1. *Direct duty cycle control* is the simplest control method. A fixed-frequency ramp is compared with the control voltage [Fig. 30.19(a)] to obtain a variable duty cycle base drive signal for the transistor. This is the simplest method of control. Disadvantages of this method are (a) provides no voltage feedforward to anticipate the effects of input voltage changes, slow response to sudden input changes, poor audio susceptibility, poor open-loop line regulation, requiring higher loop gain to achieve specifications; (b) poor dynamic response.
2. *Voltage feedforward control.* In this case the ramp amplitude varies in direct proportion to the input voltage [Fig. 30.19(b)]. The open-loop regulation is very good, and the problems in 1(a) above are corrected.
3. *Current mode control.* In this method, a second inner control loop compares the peak inductor current with the control voltage which provides improved open-loop line regulation [Fig. 30.19(c)]. All the problems of the direct duty cycle control method 1 above are corrected with this method. An additional advantage of this method is that the two-pole second-order filter is reduced to a single-pole (the filter capacitor) first-order filter, resulting in simpler compensation networks.

The above control methods can be used in all the PWM converter configurations explained below.

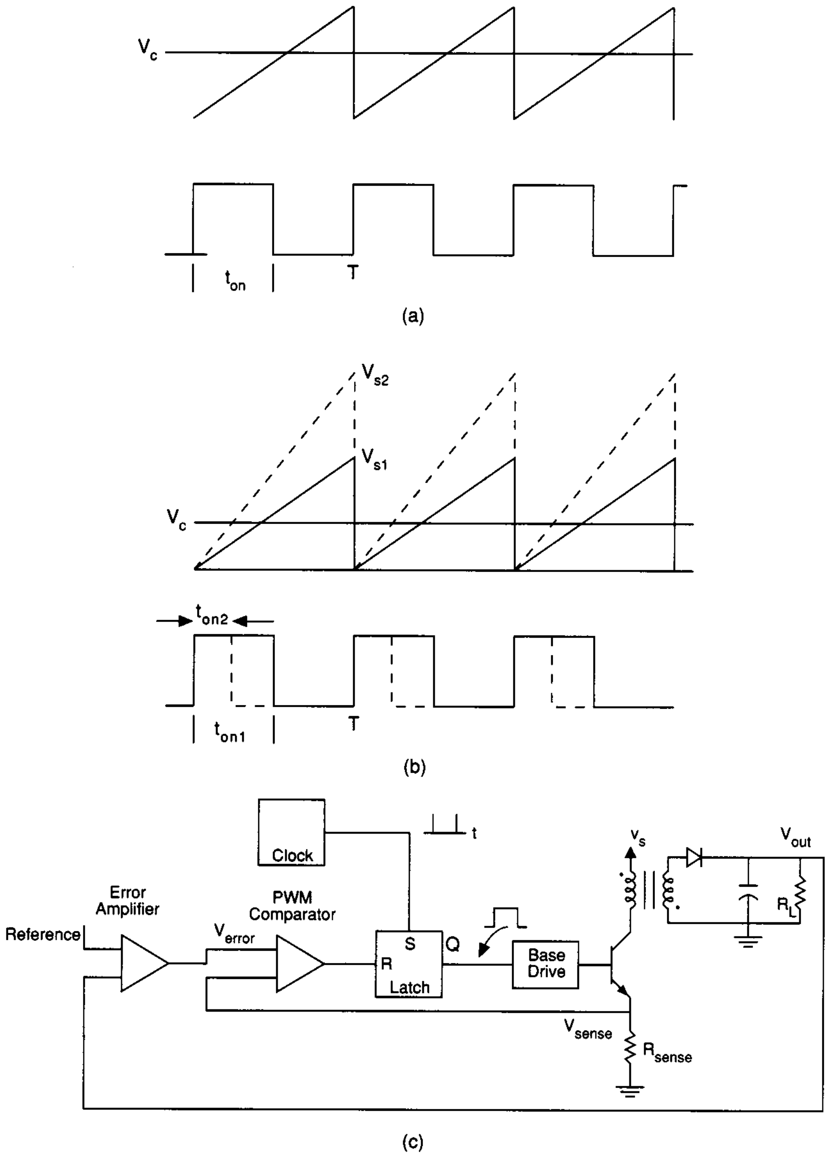


FIGURE 30.19 PWM converter control methods: (a) direct duty cycle control; (b) voltage feedforward control; (c) current mode control (illustrated for flyback converter).

PWM converters can be classified as single-ended and double-ended converters. These converters may or may not have a high-frequency transformer for isolation.

Nonisolated Single-Ended PWM Converters. The basic nonisolated single-ended converters are (a) buck (step-down), (b) boost (step-up), (c) buck-boost (step-up or step-down, also referred to as flyback), and (d) ´Cuk converters (Fig. 30.20). The first three of these converters have been discussed in Section 30.2. The ´Cuk converter provides the advantage of nonpulsating input-output current ripple requiring smaller size external filters. Output voltage expression is the same as the buck-boost converter (refer to Section 30.2) and can be less than or greater than the input voltage. There are many variations of the above basic nonisolated converters, and most of them use a high-frequency transformer for ohmic isolation between the input and the output. Some of them are discussed below.

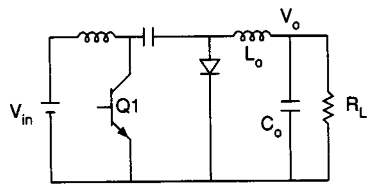


FIGURE 30.20 Nonisolated Ćuk converter.

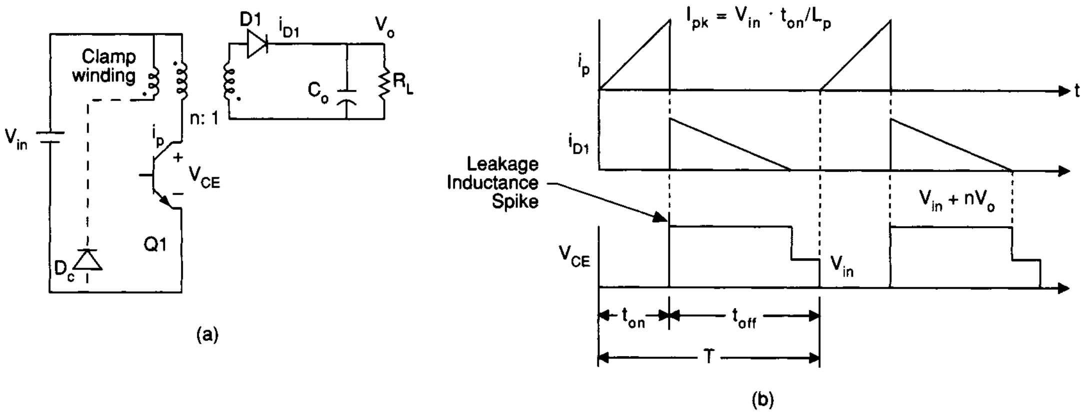


FIGURE 30.21 (a) Flyback converter. The clamp winding shown is optional and is used to clamp the transistor voltage stress to $V_{in} + nV_o$. (b) Flyback converter waveforms without the clamp winding. The leakage inductance spikes vanish with the clamp winding.

Isolated Single-Ended Topologies

1. The flyback converter (Fig. 30.21) is an **isolated** version of the buck-boost converter. In this converter (Fig. 30.21), when the transistor is on, energy is stored in the coupled inductor (not a transformer), and this energy is transferred to the load when the switch is off.

   Some of the advantages of this converter are that the leakage inductance is in series with the output diode when current is delivered to the output, and, therefore, no filter inductor is required; cross regulation for multiple output converters is good; it is ideally suited for high-voltage output applications; and it has the lowest cost.

   Some of the disadvantages are that large output filter capacitors are required to smooth the pulsating output current; inductor size is large since air gaps are to be provided; and due to stability reasons, flyback converters are usually operated in the DCM, which results in increased losses. To avoid the stability problem, flyback converters are operated with current mode control explained earlier. Flyback converters are used in the power range of 20 to 200 W.
2. The forward converter (Fig. 30.22) is based on the buck converter. It is usually operated in the CCM to reduce the peak currents and does not have the stability problem of the flyback converter. The HF transformer transfers energy directly to the output with very small stored energy. The output capacitor size and peak current rating are smaller than they are for the flyback. Reset winding is required to remove the stored energy in the transformer. Maximum duty cycle is about 0.45 and limits the control range. This topology is used for power levels up to about 1 kW.

The flyback and forward converters explained above require the rating of power transistors to be much higher than the supply voltage. The two-transistor flyback and forward converters shown in Fig. 30.23 limit the voltage rating of transistors to the supply voltage.

The Sepic converter shown in Fig. 30.24 is another isolated single-ended PWM converter.

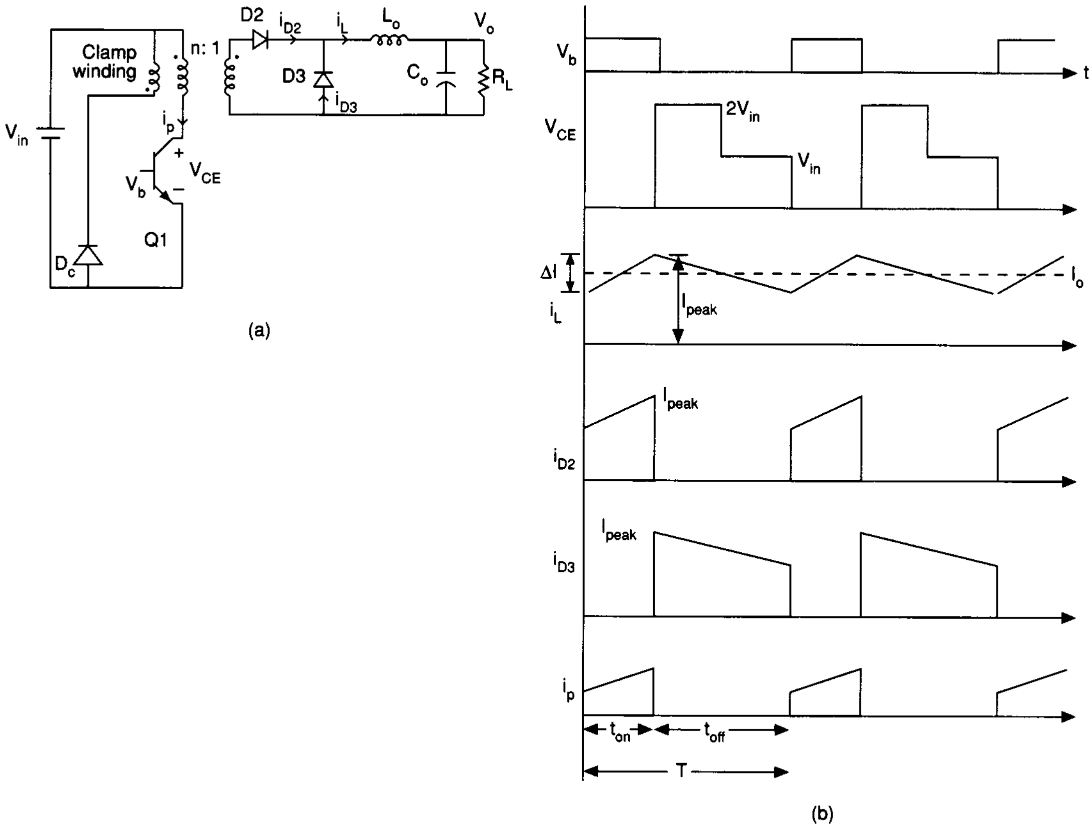


FIGURE 30.22 (a) Forward converter. The clamp winding shown is required for operation. (b) Forward converter waveforms.

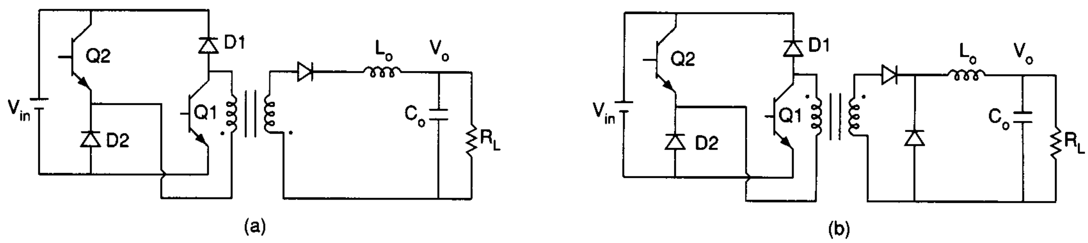


FIGURE 30.23 (a) Two-transistor single-ended flyback converter. (b) Two-transistor single-ended forward converter.

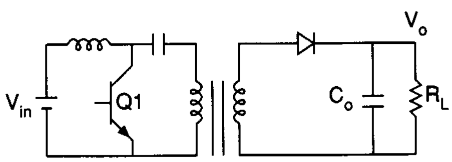


FIGURE 30.24 Sepic converter.

Double-Ended PWM Converters. Usually, for power levels above 300 W, double-ended converters are used. In double-ended converters, full-wave rectifiers are used and the output voltage ripple will have twice the switching frequency. Three important double-ended PWM converter configurations are push-pull (Fig. 30.25), half-bridge (Fig. 30.26), and full-bridge (Fig. 30.27).

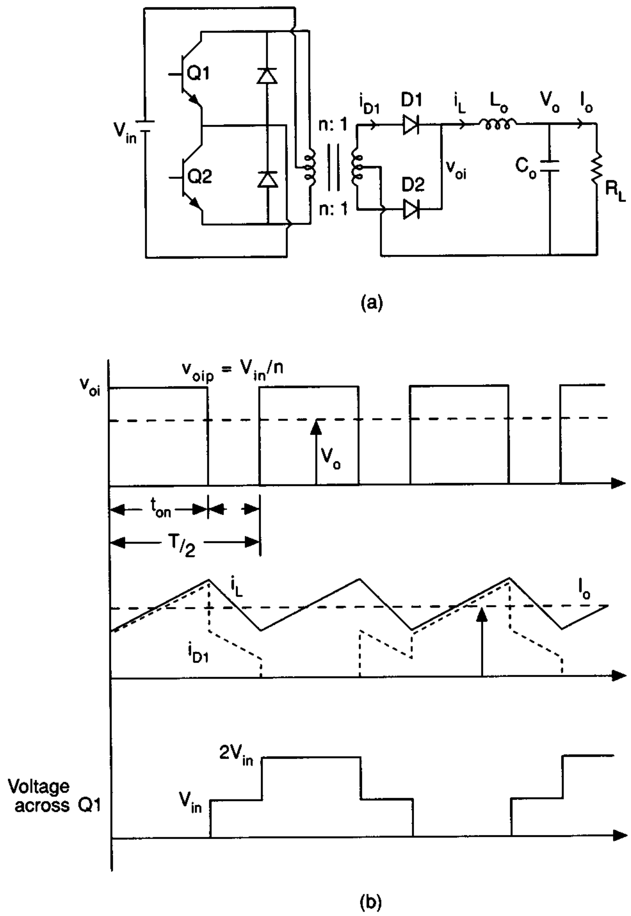


FIGURE 30.25 (a) Push-pull converter and (b) its operating waveforms.

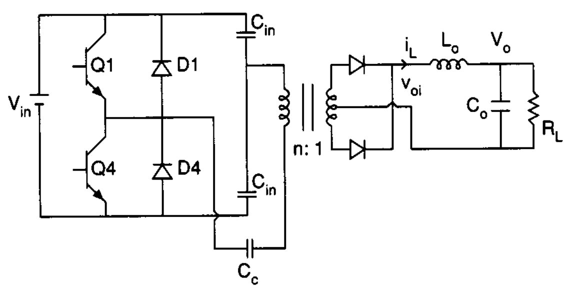


FIGURE 30.26 Half-bridge converter. Coupling capacitor C_c is used to avoid transformer saturation.

1. *The push-pull converter.* The duty ratio of each transistor in a push-pull converter (Fig. 30.25) is less than 0.5. Some of the advantages are that the transformer flux swings fully, thereby the size of the transformer is much smaller (typically half the size) than single-ended converters, and output ripple is twice the switching frequency of transistors, allowing smaller filters.

   Some of the disadvantages of this configuration are that transistors must block twice the supply voltage, flux symmetry imbalance can cause transformer saturation and special control circuitry is required to avoid this problem, and use of center-tap transformer requires extra copper resulting in higher volt-ampere (VA) rating.

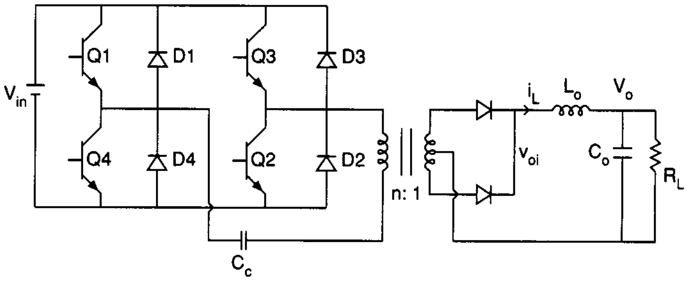


FIGURE 30.27 Full-bridge converter.

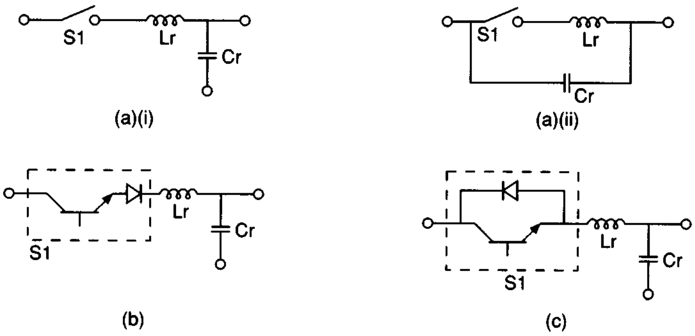


FIGURE 30.28 (a) Zero-current resonant switch: (i) L-type and (ii) M-type. (b) Half-wave configuration using L-type ZC resonant switch. (c) Full-wave configuration using L-type ZC resonant switch.

Current mode control (for the primary current) can be used to overcome the flux imbalance. This configuration is used in 100- to 500-W output range.

2. *The half-bridge.* In the half-bridge configuration (Fig. 30.26) center-tapped dc source is created by two smoothing capacitors (C_{in}), and this configuration utilizes the transformer core efficiently. The voltage across each transistor is equal to the supply voltage (half of push-pull) and, therefore, is suitable for high-voltage inputs. One salient feature of this configuration is that the input filter capacitors can be used to change between 110/220-V mains as selectable inputs to the supply.

   The disadvantage of this configuration is the requirement for large-size input filter capacitors. The half-bridge configuration is used for power levels of the order of 500 to 1000 W.

3. The full-bridge configuration (Fig. 30.27) requires only one smoothing capacitor, and for the same transistor type as that of half-bridge, output power can be doubled. It is usually used for power levels above 1 kW, and the design is more costly due to increased number of components (uses four transistors compared to two in push-pull and half-bridge converters).

   One of the salient features of a full-bridge converter is that by using proper control technique it can be operated in zero-voltage switching (ZVS) mode. This type of operation results in negligible switching losses. However, at reduced load currents, the ZVS property is lost. Recently, there has been a lot of effort to overcome this problem.

Resonant Power Supplies

Similar to the PWM converters, there are two types of resonant converters: single-ended and double-ended. Resonant converter configurations are obtained from the PWM converters explained earlier by adding LC (inductor-capacitor) resonating elements to obtain sinusoidally varying voltage and/or current waveforms. This approach reduces the switching losses and the switch stresses during switching instants, enabling the converter to operate at high switching frequencies, resulting in reduced size, weight, and cost. Some other advantages of resonant converters are that leakage inductances of HF transformers and the junction capacitances of semiconductors can

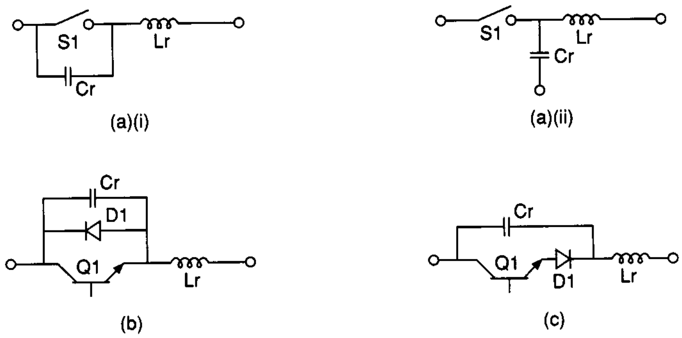


FIGURE 30.29 (a) Zero-voltage resonant switches. (b) Half-wave configuration using ZV resonant switch shown in Fig. (a)(i). (c) Full-wave configuration using ZV resonant switch shown in Fig. (a)(i).

be used profitably in the resonant circuit, and reduced EMI. The major disadvantage of resonant converters is increased peak current (or voltage) stress.

Single-Ended Resonant Converters. They are referred to as quasi-resonant converters (QRCs) since the voltage (or current) waveforms are quasi-sinusoidal in nature. The QRCs can operate with zero-current switching (ZCS) or ZVS or both. All the QRC configurations can be generated by replacing the conventional switches by the resonant switches shown in Figs. 30.28 and 30.29. A number of configurations are realizable. Basic principles of ZCS and ZVS are explained briefly below.

1. *Zero-current switching QRCs* [Sum, 1988; Liu et al., 1985]. Figure 30.30(a) shows an example of a ZCS QR buck converter implemented using a ZC resonant switch. Depending on whether the resonant switch is half-wave type or full-wave type, the resonating current will be only half-wave sinusoidal [Fig. 30.30(b)] or a full sine-wave [Fig. 30.30(c)]. The device currents are shaped sinusoidally, and, therefore, the switching losses are almost negligible with low turn-on and turn-off stresses. ZCS QRCs can operate at frequencies of the order of 2 MHz. The major problems with this type of converter are high peak currents through the switch and capacitive turn-on losses.
2. *Zero-voltage switching QRCs* [Sum, 1988; Liu and Lee, 1986]. ZVS QRCs are duals of ZCS QRCs. The auxiliary LC elements are used to shape the switching device's voltage waveform at off time in order to create a zero-voltage condition for the device to turn on. Fig. 30.31(a) shows an example of ZVS QR boost converter implemented using a ZV resonant switch. The circuit can operate in the half-wave mode [Fig. 30.31(b)] or in the full-wave mode [Fig. 30.31(c)] depending on whether a half-wave or full-wave ZV resonant switch is used, and the name comes from the capacitor voltage waveform. The full-wave mode ZVS circuit suffers from capacitive turn-on losses. The ZVS QRCs suffer from increased voltage stress on the switch. However, they can be operated at much higher frequencies compared to ZCS QRCs.

Double-Ended Resonant Converters. These converters [Sum, 1988; Bhat, 1991; Steigerwald, 1988; Bhat, 1992] use full-wave rectifiers at the output, and they are generally referred to as resonant converters. A number of resonant converter configurations are realizable by using different resonant tank circuits, and the three most popular configurations, namely, the series resonant converter (SRC), the parallel resonant converter (PRC), and the series-parallel resonant converter (SPRC) (also called LCC-type PRC), are shown in Fig. 30.32.

Series resonant converters [Fig. 30.32(a)] have high efficiency from full load to part load. Transformer saturation is avoided due to the series blocking resonating capacitor. The major problems with the SRC are that it requires a very wide change in switching frequency to regulate the load voltage and the output filter capacitor must carry high ripple current (a major problem especially in low output voltage, high output current applications).

Parallel resonant converters [Fig. 30.32(b)] are suitable for low output voltage, high output current applications due to the use of filter inductance at the output with low ripple current requirements for the filter capacitor. The major disadvantage of the PRC is that the device currents do not decrease with the load current, resulting in reduced efficiency at reduced load currents.

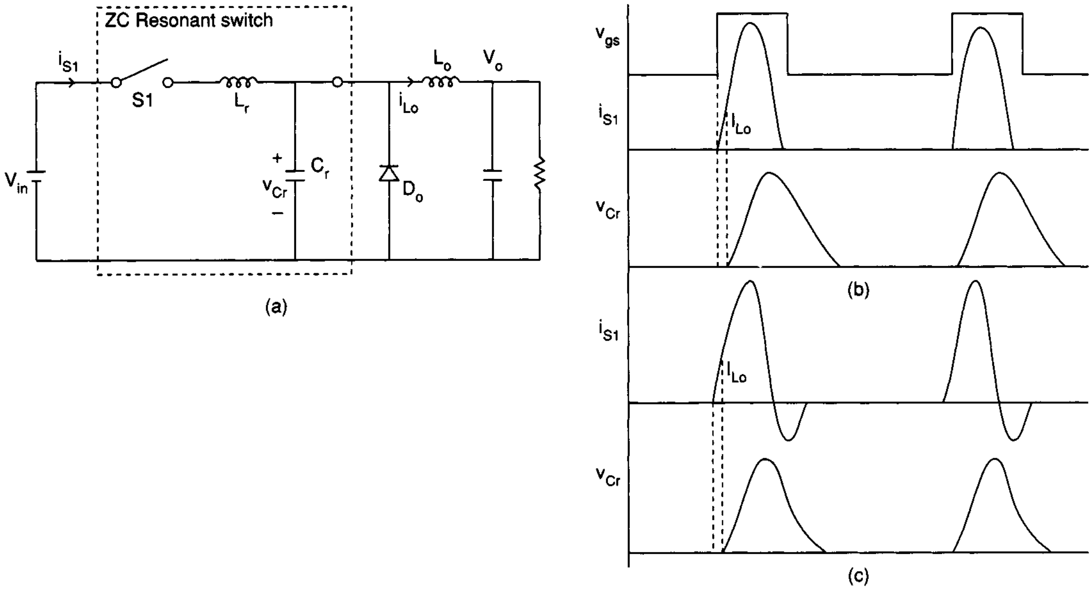


FIGURE 30.30 (a) Implementation of ZCS QR buck converter using L-type resonant switch. (b) Operating waveforms for half-wave mode. (c) Operating waveforms for full-wave mode.

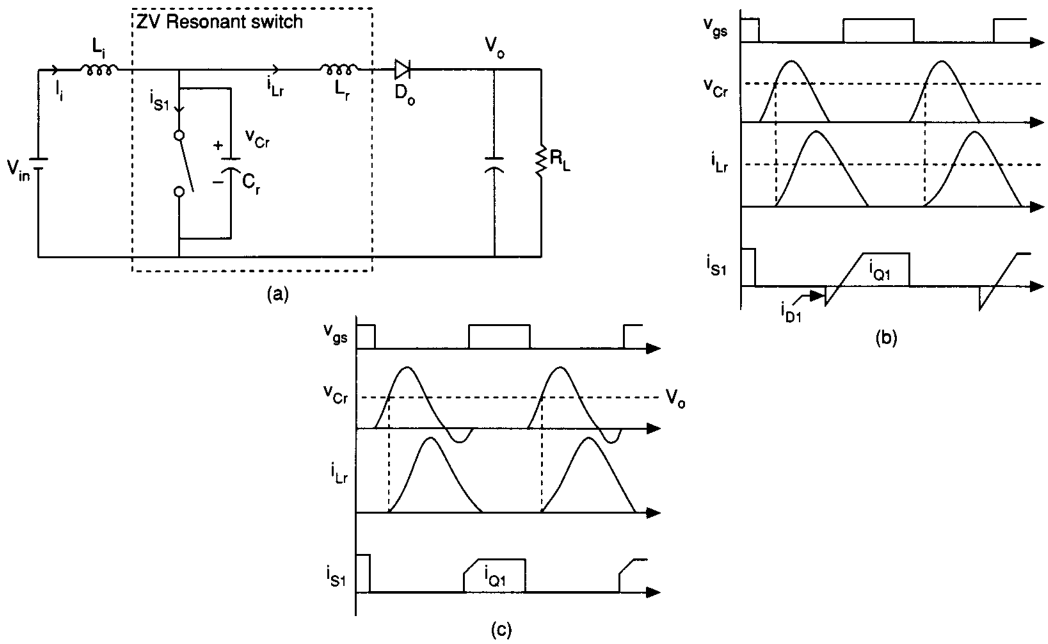


FIGURE 30.31 (a) Implementation of ZVS QR buck converter using resonant switch shown in Fig. 30.28(a)(i). (b) Operating waveforms for half-wave mode. (c) Operating waveforms for full-wave mode.

The SPRC [Fig. 30.32(c)] takes the desirable features of SRC and PRC.

Load voltage regulation in resonant converters for input supply variations and load changes is achieved by either varying the switching frequency or using fixed-frequency (variable pulsewidth) control.

1. *Variable-frequency operation.* Depending on whether the switching frequency is below or above the natural resonance frequency (ω_r), the converter can operate in different operating modes as explained below.

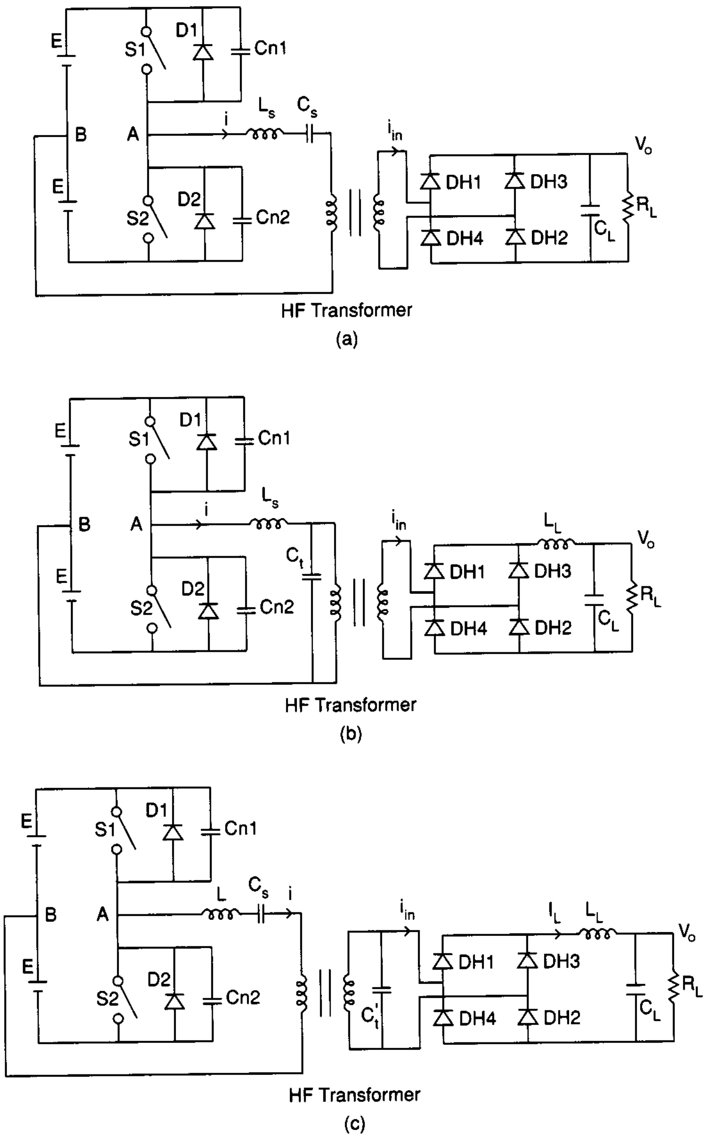


FIGURE 30.32 High-frequency resonant converter (half-bridge version) configurations suitable for operation above resonance. *Cn*1 and *Cn*2 are the snubber capacitors. (*Note:* For operation below resonance, *di/dt* limiting inductors and RC snubbers are required. For operation above resonance, only capacitive snubbers are required as shown.) (a) Series resonant converter. Leakage inductances of the HF transformer can be part of resonant inductance. (b) Parallel resonant converter. (c) Series-parallel (or LCC-type) resonant converter with capacitor C_t placed on the secondary side of the HF transformer.

a. Below-resonance (leading PF) mode. When the switching frequency is below the natural resonance frequency, the converter operates in a below-resonance mode (Fig. 30.33). The equivalent impedance across *AB* presents a leading PF so that natural turn-off of the switches is assured and any type of fast turn-off switch (including asymmetric SCRs) can be used. Depending on the instant of turn-on of switches S_1 and S_2, the converter can enter into two modes of operation, namely, continuous and discontinuous current modes. The steady-state operation in continuous current mode (CCM) [Fig. 30.33(a)] is explained briefly as follows.

Assume that diode D_2 was conducting and switch S_1 is turned on. The current carried by D_2 will be transferred to S_1 almost instantaneously (except for a small time of recovery of D_2 during which input supply is shorted through D_2 and S_1, and the current is limited by the *di/dt* limiting inductors). The

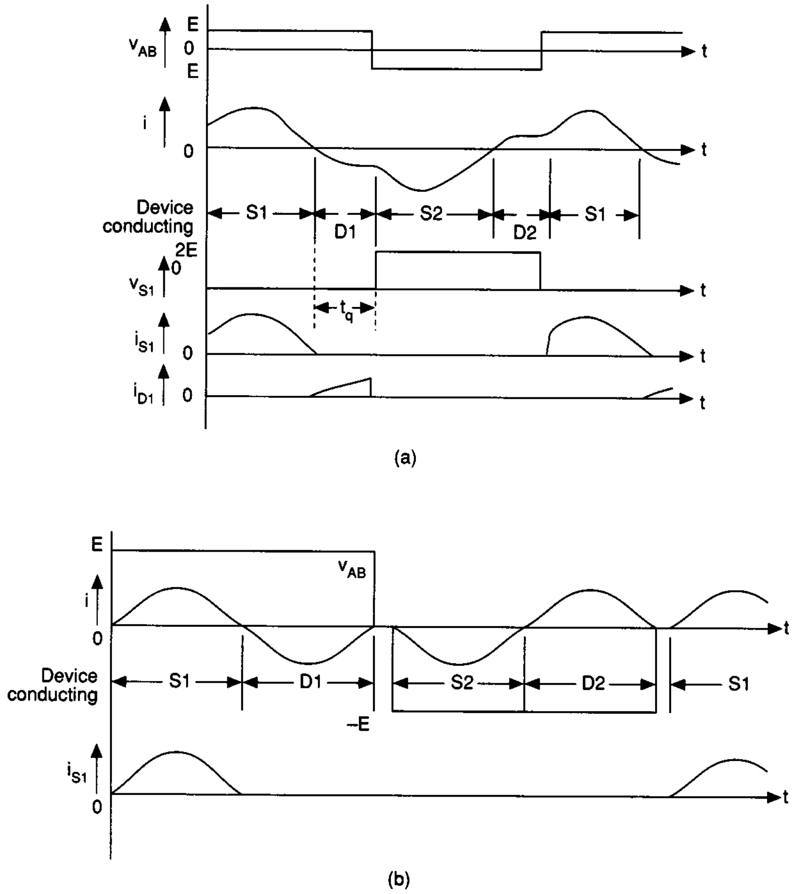


FIGURE 30.33 Typical waveforms at different points of a resonant converter operating below resonance (a) in continuous current mode and (b) in discontinuous current mode.

current i then oscillates sinusoidally and goes to zero in the natural way. The current tries to reverse, and the path for this current is provided by the diode D_1. Conduction of D_1 feeds the reactive energy in the load and the tank circuit back to the supply. The on-state of D_1 also provides a reverse voltage across S_1, allowing it to turn off. After providing a time equal to or greater than the turn-off time of S_1, switch S_2 can be turned on to initiate the second half cycle. The process is similar to the first half cycle, with the voltage across v_{AB} being of opposite polarity, and the functions of D_1, S_1 will be assumed by D_2, S_2. With this type of operation, the converter works in the continuous current mode as the switches are turned on before the currents in the diodes reach zero. If the switching on of S_1 and S_2 is delayed such that the currents through the previously conducting diodes reach zero, then there are zero current intervals and the **inverter** operates in the DCM [Fig. 30.33(b)].

Load voltage regulation is achieved by decreasing the switching frequency below the rated value. Since the inverter output current i leads the inverter output voltage v_{AB}, this type of operation is also called a leading PF mode of operation. If transistors are used as the switching devices, then for operation in DCM, the pulsewidth can be kept constant while decreasing the switching frequency to avoid CCM operation. DCM operation has the advantages of negligible switching losses due to ZCS, lower *di/dt* and *dv/dt* stresses, and simple control circuitry. However, DCM operation results in higher switch peak currents.

From the waveforms shown in Fig. 30.33, the following problems can be identified for operation in the below-resonance mode: requirement of *di/dt* inductors to limit the large turn-on switch currents and a need for lossy RC snubbers and fast recovery diodes. Since the switching frequency is decreased to control the load power, the HF transformer and magnetics must be designed for the lowest switching frequency, resulting in increased size of the converter.

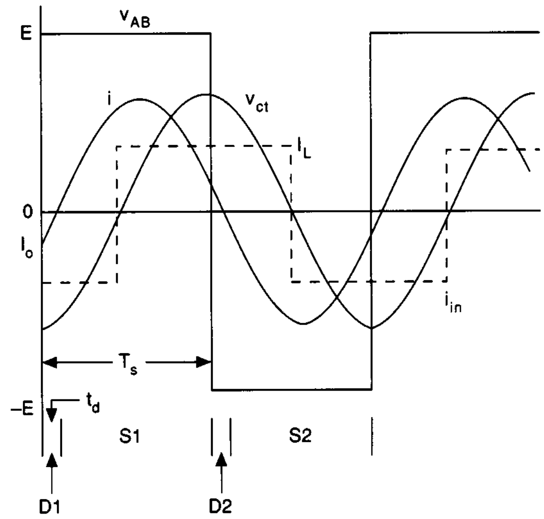


FIGURE 30.34 Typical operating waveforms at different points of an SPRC operating above resonance.

b. Above-resonance (lagging PF) mode. If switches capable of gate or base turn-off (e.g., MOSFETs, bipolar transistors) are used, then the converter can operate in the above-resonance mode (lagging PF mode). Figure 30.34 shows some typical operating waveforms for such type of operation, and it can be noticed that the current i lags the voltage v_{AB}. Since the switch takes current from its own diode across it at zero-current point, there is no need for *di/dt* limiting inductance, and a simple capacitive snubber can be used. In addition, the internal diodes of MOSFETs can be used due to the large turn-off time available for the diodes. Major problems with the lagging PF mode of operation are that there are switch turn-off losses, and since the voltage regulation is achieved by increasing the switching frequency above the rated value, the magnetic losses increase and the design of a control circuit is difficult.

Exact analysis of resonant converters is complex due to the nonlinear loading on the resonant tanks. The rectifier-filter-load resistor block can be replaced by a square-wave voltage source [for SRC, Fig. 30.32(a)] or a square-wave current source [for PRC and SPRC, Fig. 30.32(b) and (c)]. Using fundamental components of the waveforms, an approximate analysis [Bhat, 1991; Steigerwald, 1988] using a phasor circuit gives a reasonably good design approach. This analysis approach is illustrated next for the SPRC.

2. *Approximate analysis of SPRC.* Figure 30.35 shows the equivalent circuit at the output of the inverter and the phasor circuit used for the analysis. All the equations are normalized using the base quantities

$$\text{Base voltage } V_B = E_{\min}$$

$$\text{Base impedance } Z_B = R'_L = n^2 R_L$$

$$\text{Base current } I_B = V_B / I_B$$

The converter gain [normalized output voltage in per unit (p.u.) referred to the primary-side] can be derived as [Bhat, 1991; Steigerwald, 1988]

$$M = \frac{1}{\left\{\left(\frac{\pi^2}{8}\right)^2\left[1 + \left(\frac{C_t}{C_s}\right)\left(1 - y_s^2\right)\right]^2 + Q_s^2\left[y_s - \left(\frac{1}{y_s}\right)\right]^2\right\}^{1/2}} \quad \text{p.u.} \tag{30.1}$$

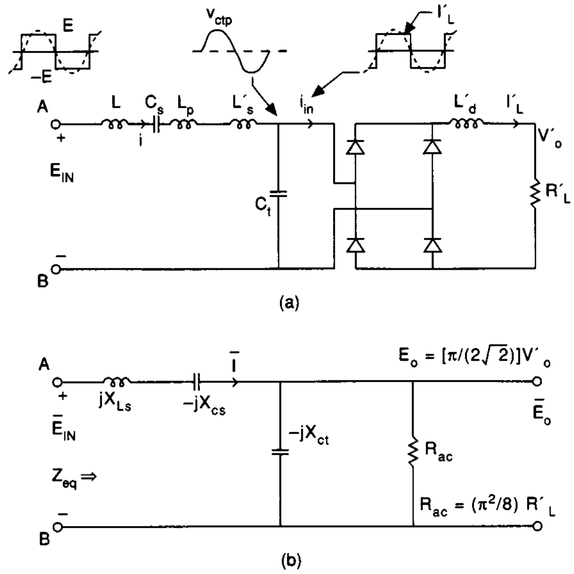


FIGURE 30.35 (a) Equivalent circuit for a SPRC at the output of the inverter terminals (across *AB*) of Fig. 30.31(c), L_p and L'_s are the leakage inductance of the primary and primary referred leakage inductance of the secondary, respectively. (b) Phasor circuit model used for the analysis of the SPRC converter.

where

$$Q_s = \frac{(L_s / C_s)^{1/2}}{R'_L}\ ; L_s = L + L_p + L'_s \tag{30.2}$$

$$y_s = \frac{f_s}{f_r} \tag{30.3}$$

and

$$\begin{aligned} f_s &= \text{switching frequency} \\ f_r &= \text{series resonance frequency} \\ &= \frac{\omega_r}{2\pi} = \frac{1}{2\pi(L_s C_s)^{1/2}} \end{aligned} \tag{30.4}$$

The equivalent impedance looking into the terminals *AB* is given by

$$Z_{eq} = \frac{B_1 + jB_2}{B_3} \quad \text{p.u.} \tag{30.5}$$

where

$$B_1 = \left(\frac{8}{\pi^2}\right)\left(\frac{C_s}{C_t}\right)^2\left(\frac{Q_s}{y_s}\right)^2 \tag{30.6}$$

$$B_2 = Q_s\left(y_s - \frac{1}{y_s}\right)\left[1 + \left(\frac{8}{\pi^2}\right)^2\left(\frac{C_s}{C_t}\right)^2\left(\frac{Q_s}{y_s}\right)^2\right] - \left(\frac{C_s}{C_t}\right)\left(\frac{Q_s}{y_s}\right) \tag{30.7}$$

$$B_3 = 1 + \left(\frac{8}{\pi^2}\right)^2\left(\frac{C_s}{C_t}\right)^2\left(\frac{Q_s}{y_s}\right)^2 \tag{30.8}$$

The peak inverter output (resonant inductor) current can be calculated using

$$I_p = \frac{4}{\pi\left|Z_{eq}\right|} \quad \text{p.u.} \tag{30.9}$$

The same current flows through the switching devices.

The value of initial current I_0 is given by

$$I_0 = I_p \sin(-\phi) \text{ p.u.} \tag{30.10}$$

where $\phi = \tan^{-1}(B_2/B_1)$ rad. B_1 and B_2 are given by Eqs. (30.6) and (30.7), respectively.

If I_0 is negative, then forced commutation is necessary and the converter is operating in the lagging PF mode. The peak voltage across the capacitor C'_t (on the secondary side) is

$$V_{\text{ctp}} = \frac{\pi}{2} V_o \quad \text{V} \tag{30.11}$$

The peak voltage across C_s and the peak current through C'_t are given by

$$V_{\text{csp}} = \frac{Q_s}{y_s} I_p \quad \text{p.u.} \tag{30.12}$$

$$I_{\text{ctp}} = \frac{V_{\text{ctp}}}{X_{\text{cptu}} R_L} \quad \text{A} \tag{30.13}$$

$$X_{\text{ctpu}} = \left(\frac{C_s}{C_t}\right)\left(\frac{Q_s}{y_s}\right) \quad \text{p.u.} \tag{30.14}$$

The plot of converter gain versus the switching frequency ratio y_s, obtained using (30.1), is shown for $C_s/C_t = 1$ in Fig. 30.36, for the lagging PF mode of operation. If the ratio C_s/C_t increases, then the converter takes the characteristics of SRC and the load voltage regulation requires a very wide range in the frequency change. Lower values of C_s/C_t take the characteristics of a PRC. Therefore, a compromised value of $C_s/C_t = 1$ is chosen.

It is possible to realize higher-order resonant converters with improved characteristics and many of them are presented in Bhat [1991].

3. *Fixed-frequency operation.* To overcome some of the problems associated with the variable frequency control of resonant converters, they are operated with fixed frequency [Sum, 1988; Bhat, 1992]. A number of configurations and control methods for fixed-frequency operation are available in the literature (Bhat [1992] gives a list of papers). One of the most popular methods of control is the phase-shift control

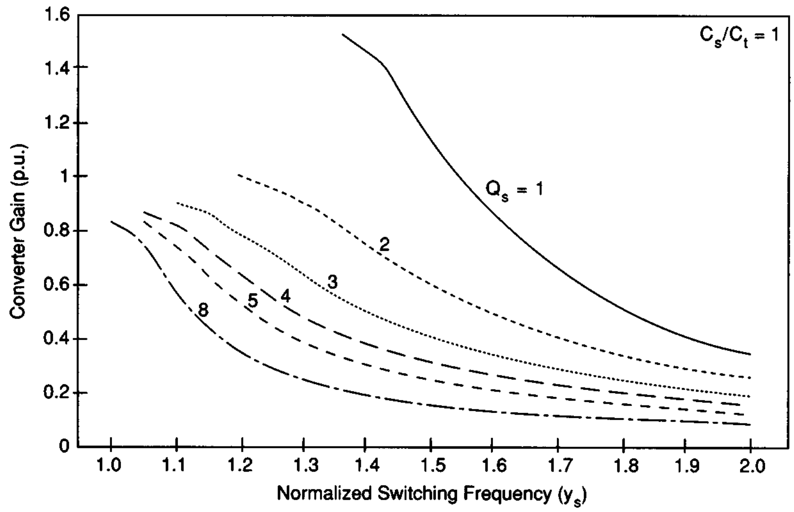


FIGURE 30.36 The converter gain M (p.u.) (normalized output voltage) versus normalized switching frequency y_s of SPRC operating above resonance for $C_s/C_t = 1$.

(also called clamped-mode or PWM operation) method. Figure 30.37 illustrates the clamped-mode fixed-frequency operation of the SPRC. The load power control is achieved by changing the phase-shift angle ϕ between the gating signals to vary the pulsewidth of v_{AB}.

4. *Design example.* Design a 500-W output SPRC (half-bridge version) with secondary-side resonance (operation in lagging PF mode and variable-frequency control) with the following specifications:

$$\text{Minimum input supply voltage} = 2E_{\min} = 230 \text{ V}$$

$$\text{Load voltage, } V_o = 48 \text{ V}$$

$$\text{Switching frequency, } f_s = 100 \text{ kHz}$$

$$\text{Maximum load current} = 10.42 \text{ A}$$

As explained in item 2, $C_s/C_t = 1$ is chosen. Using the constraints (1) minimum kVA rating of tank circuit per kW output power, (2) minimum inverter output peak current, and (3) enough turn-off time for the switches, it can be shown that [Bhat, 1991] $Q_s = 4$ and $y_s = 1.1$ satisfy the design constraints. From Fig. 30.36, $M = 0.8$ p.u.

Average load voltage referred to the primary side of the HF transformer = 0.8 × 115 V = 92 V. Therefore, the transformer turns ratio required ≃1.84.

$$R_L' = n^2\left(\frac{V_o^2}{P_o}\right) = 15.6\ \Omega$$

The values of L_s and C_s can be obtained by solving

$$\left(\frac{L_s}{C_s}\right)^{1/2} = 4 \times 15.6\,\Omega \text{ and } \omega_r = \frac{1}{(L_s C_s)^{1/2}} = 2\pi\frac{f_s}{y_s}$$

Solving the above equations gives $L_s = 109$ μH and $C_s = 0.0281$ μF. Leakage inductance ($L_p + L'_s$) of the HF transformer can be used as part of L_s. Typical value for a 100-kHz practical transformer (using Tokin

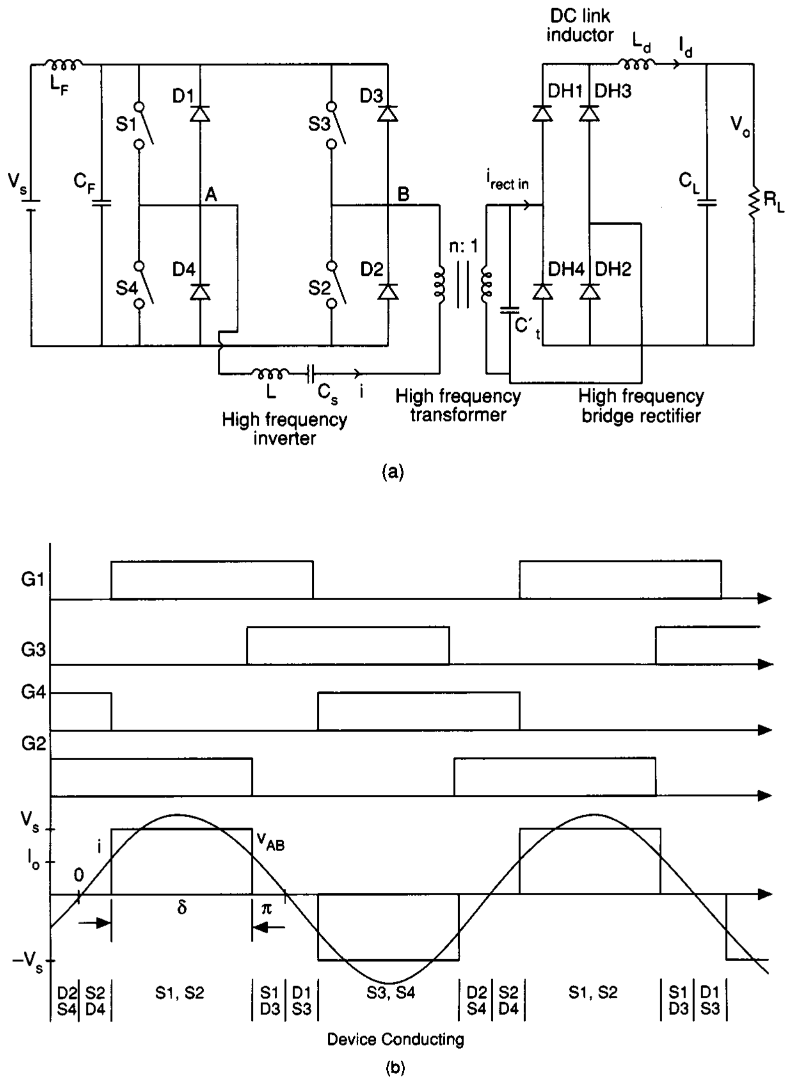


FIGURE 30.37 (a) Basic circuit diagram of series-parallel resonant converter suitable for fixed-frequency operation with PWM (clamped-mode) control. (b) Waveforms to illustrate the operation of fixed-frequency PWM series-parallel resonant converter working with a pulsewidth δ.

Mn-Zn 2500B2 Ferrite, E-I type core) for this application is about 5 μH. Therefore, the external resonant inductance required is $L = 104$ μH.

Since $C_s/C_t = 1$ is chosen, $C_t = 0.0281$ μF. The actual value of C_t used on the secondary side of the HF transformer = $(1.84)^2 \times 0.0281 = 0.09514$ μF. The resonating capacitors must be HF type (e.g., polypropylene) and must be capable of withstanding the voltage and current ratings obtained above (enough safety margin must be provided).

Using Eqs. (30.9) and (30.11) to (30.13):

$$\text{Peak current through switches} = 7.6 \text{ A}$$

$$\text{Peak voltage across } C_s,\ V_{csp} = 430 \text{ V}$$

$$\text{Peak voltage (on secondary side) across } C'_t,\ V_{ctp} = 76 \text{ V}$$

$$\text{Peak current through capacitor } C'_t \text{ (on secondary side)},\ I_{ctp} = 4.54 \text{ A}$$

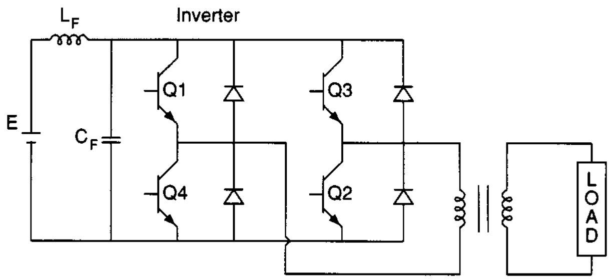


FIGURE 30.38 An inverter circuit to obtain variable-voltage, variable-frequency ac source. Using sinusoidal pulsewidth modulation control scheme, sine-wave ac output voltage can be obtained.

A simple control circuit can be built using PWM IC SG3525 and TSC429 MOSFET driver ICs.

With the development of digital ICs operating on low-voltage (of the order of 3 V) supplies, use of MOSFETs as *synchronous rectifiers* with very low voltage drop (~0.2 V) has become essential [Motorola, 1989] to increase the efficiency of the power supply.

AC Power Supplies

Some applications of ac power supplies are ac motor drives, **uninterruptible power supply (UPS)** used as a standby ac source for critical loads (e.g., in hospitals, computers), and dc source-to-utility interface (either to meet peak power demands or to augment energy by connecting unconventional energy sources like photovoltaic arrays to the utility line). In ac induction motor drives, the ac power main is rectified and filtered to obtain a smooth dc source, and then an inverter (single-phase version is shown in Fig. 30.38) is used to obtain a variable-frequency, variable-voltage ac source. The sinusoidal pulsewidth modulation technique described in Section 30.2 can be used to obtain a sinusoidal output voltage. Some other methods used to get sinusoidal voltage output are [Rashid, 1988] a number of phase-shifted inverter outputs summed in an output transformer to get a stepped waveform that approximates a sine wave and the use of a bang-bang controller in Fig. 30.38. All these methods use line-frequency (60 Hz) transformers for voltage translation and isolation purposes. To reduce the size, weight, and cost of such systems, one can use dc-to-dc converters (discussed earlier) as an intermediate stage. Figure 30.39 shows such a system in block schematic form. One can use an HF inverter circuit (discussed earlier) followed by a **cycloconverter** stage. The major problem with these schemes is the reduction in efficiency due to the extra power stage. Figure 30.40 shows a typical UPS scheme. The battery shown has to be charged by a separate rectifier circuit.

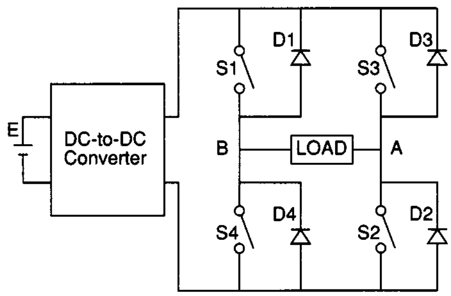


FIGURE 30.39 AC power supplies using HF switching (PWM or resonant) dc-to-dc converter as an input stage. HF transformer isolated dc-to-dc converters can be used to reduce the size and weight of the power supply. Sinusoidal voltage output can be obtained using the modulation in the output inverter stage or in the dc-to-dc converter.

AC-to-ac conversion can also be achieved using cycloconverters [e.g., Rashid, 1988].

Special Power Supplies

Using the inverters and cycloconverters, it is possible to realize bidirectional ac and dc power supplies. In these power supplies [Rashid, 1988], power can flow in both directions, i.e., from input to output or from output to input. It is also possible to control the ac-to-dc converters to obtain sinusoidal line current with unity PF and low harmonic distortion at the ac source.

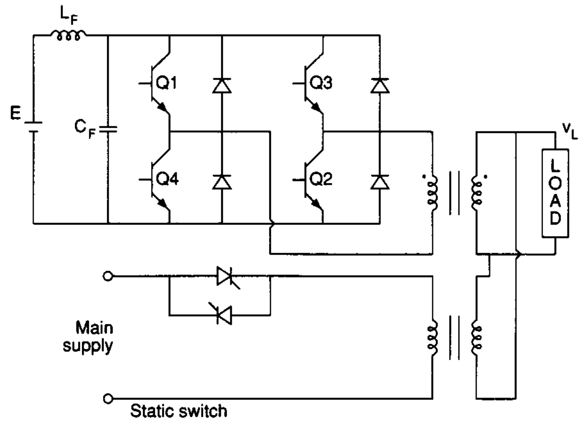


FIGURE 30.40 A typical arrangement of UPS system. The load gets power through the static switch when the ac main supply is present. The inverter supplies power when the main supply fails.

Defining Terms

Converter: A circuit that performs one of the following power conversions — ac to dc, dc to dc, dc to ac, or ac to ac.

Cycloconverter: A power electronic circuit that converts ac input to ac output (generally) of lower frequency than the input source without using any intermediate dc state.

Inverter: A power electronic circuit that converts dc input to ac output.

Isolated: A power electronic circuit that has ohmic isolation between the input source and the load circuit.

Pulsewidth-modulated (PWM) converters: A power electronic converter that employs square-wave switching waveforms with variation of pulsewidth for controlling the load voltage.

Regulated output: Output load voltage is kept at the required value for changes in either the load or the input supply voltage.

Resonant converters: A power electronic converter that employs "LC resonant circuits" to obtain sinusoidal switching waveforms.

Uninterruptible power supply (UPS): A stand-by dc-to-ac inverter used mostly to provide an emergency power to loads at mains frequency (50/60 Hz) in the event of a mains failure.

References

A.K.S. Bhat, "A unified approach for the steady-state analysis of resonant converters," *IEEE Trans. Industrial Electronics,* vol. 38, no. 4, pp. 251–259, Aug. 1991.

A.K.S. Bhat, "Fixed frequency PWM series-parallel resonant converter," *IEEE Trans. Industry Applications*, vol. 28, no. 5, pp. 1002–1009, 1992.

E.R. Hnatek, *Design of Solid-State Power Supplies*, 2nd ed., New York: Van Nostrand Reinhold, 1981.

K.H. Liu and F.C. Lee, "Zero-Voltage Switching Technique In DC/DC Converters," IEEE Power Electronics Specialists Conference Record, 1986, pp. 58–70.

K.H. Liu, R. Oruganti, and F.C. Lee, "Resonant Switches—Topologies and Characteristics," IEEE Power Electronics Specialists Conference Record, 1985, pp. 106–116.

Motorola, *Linear/Switchmode Voltage Regulator Handbook*, 1989.

Philips Semiconductors, *Power Semiconductor Applications*, 1991.

M.H. Rashid, *Power Electronics: Circuits, Devices, and Applications*, Englewood Cliffs, N.J.: Prentice-Hall, 1988.

R. Severns and G. Bloom, *Modern Switching DC-to-DC Converters*, New York: Van Nostrand Reinhold, 1988.

R.L. Steigerwald, "A comparison of half-bridge resonant converter topologies," *IEEE Trans. Power Electron.,* vol. PE-3, no. 2, pp. 174–182, April 1988.

K.K. Sum, *Recent Developments in Resonant Power Conversion*, Calif.: Intertech Communications, 1988.

Unitrode Switching Regulated Power Supply Design Seminar Manual, Lexington, Mass.: Unitrode Corporation, 1984.

Further Information

The following monthly magazines and conference records publish papers on the analysis, design, and experimental aspects of power supply configurations and their applications:

IEEE Transactions on Power Electronics, IEEE Transactions on Industrial Electronics, IEEE Transactions on Industry Applications, and *IEEE Transactions on Aerospace and Electronic Systems.*

IEEE Power Electronics Specialists Conference Records, IEEE Applied Power Electronics Conference Records, IEEE Industry Applications Conference Records, and *IEEE International Telecommunications Energy Conference Records.*

30.4 Converter Control of Machines

Bimal K. Bose

Converter-controlled electrical machine drives are very important in modern industrial applications. Some examples in the high-power range are metal rolling mills, cement mills, and gas line compressors. In the medium-power range are textile mills, paper mills, and subway car propulsion. Machine tools and computer peripherals are examples of converter-controlled electrical machine drive applications in the low-power range. The converter normally provides a variable-voltage dc power source for a dc motor drive and a variable-frequency, variable-voltage ac power source for an ac motor drive. The drive system efficiency is high because the converter operates in switching mode using power semiconductor devices. The primary control variable of the machine may be torque, speed, or position, or the converter can operate as a solid-state starter of the machine. The recent evolution of high-frequency power semiconductor devices and high-density and economical microelectronic chips, coupled with converter and control technology developments, is providing a tremendous boost in the applications of drives.

Converter Control of DC Machines

The speed of a dc motor can be controlled by controlling the dc voltage across its armature terminals. A phase-controlled thyristor converter can provide this dc voltage source. For a low-power drive, a single-phase bridge converter can be used, whereas for a high-power drive, a three-phase bridge circuit is preferred. The machine can be a permanent magnet or wound field type. The wound field type permits variation and reversal of field and is normally preferred in large power machines.

Phase-Controlled Converter DC Drive

Figure 30.41 shows a dc drive using a three-phase thyristor bridge converter. The converter rectifies line ac voltage to variable dc output voltage by controlling the firing angle of the thyristors. With rated field excitation, as the armature voltage is increased, the machine will develop speed in the forward direction until the rated, or base, speed is developed at full voltage when the firing angle is zero. The motor speed can be increased further by weakening the field excitation. Below the base speed, the machine is said to operate in constant

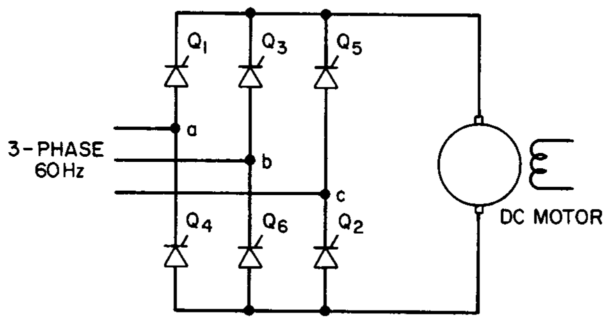


FIGURE 30.41 Three-phase thyristor bridge converter control of a dc machine.

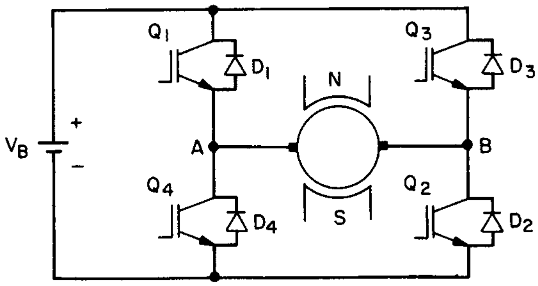


FIGURE 30.42 Four-quadrant dc motor drive using an H-bridge converter.

torque region, whereas the field weakening mode is defined as the constant power region. At any operating speed, the field can be reversed and the converter firing angle can be controlled beyond 90 degrees for **regenerative braking** mode operation of the drive. In this mode, the motor acts as a generator (with negative induced voltage) and the converter acts as an inverter so that the mechanical energy stored in the inertia is converted to electrical energy and pumped back to the source. Such **two-quadrant** operation gives improved efficiency if the drive accelerates and decelerates frequently. The speed of the machine can be controlled with precision by a feedback loop where the command speed is compared with the machine speed measured by a tachometer. The speed loop error generally generates the armature current command through a compensator. The current is then feedback controlled with the firing angle control in the inner loop. Since torque is proportional to armature current (with fixed field), a current loop provides direct torque control, and the drive can accelerate or decelerate with the rated torque. A second bridge converter can be connected in antiparallel so that the dual converter can control the machine speed in all the four quadrants (motoring and regeneration in forward and reverse speeds).

Pulsewidth Modulation Converter DC Machine Drive

Four-quadrant speed control of a dc drive is also possible using an H-bridge pulsewidth modulation (PWM) converter as shown in Fig. 30.42. Such drives (using a permanent magnet dc motor) are popular in low-power applications, such as robotic and instrumentation drives. The dc source can be a battery or may be obtained from ac supply through a diode rectifier and filter. With PWM operation, the drive response is very fast and the armature current ripple is small, giving less harmonic heating and torque pulsation. Four-quadrant operation can be summarized as follows:

Quadrant 1: Forward motoring (buck or step-down converter mode)
- Q_1—on
- Q_3, Q_4—off
- Q_2—chopping
- Current freewheeling through D_3 and Q_1

Quadrant 2: Forward regeneration (boost or step-up converter mode)
- Q_1, Q_2, Q_3—off
- Q_4—chopping
- Current freewheeling through D_1 and D_2

Quadrant 3: Reverse motoring (buck converter mode)
- Q_3—on
- Q_1, Q_2—off
- Q_4—chopping
- Current freewheeling through D_1 and Q_3

Quadrant 4: Reverse regeneration (boost converter mode)
- Q_1, Q_3, Q_4—off
- Q_2—chopping
- Current freewheeling through D_3 and D_4

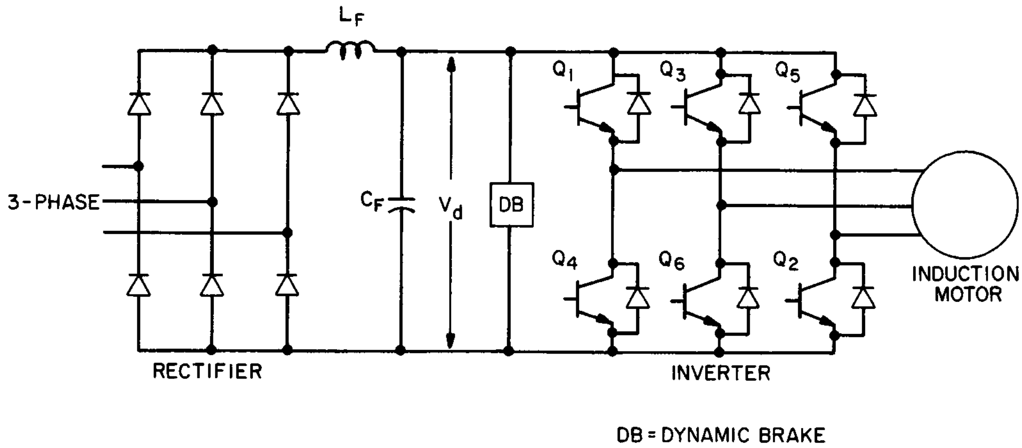


FIGURE 30.43 Diode rectifier PWM inverter control of an induction motor.

Often a drive may need only a one- or two-quadrant mode of operation. In such a case, the converter topology can be simple. For example, in one-quadrant drive, only Q_2 chopping and D_3 freewheeling devices are required, and the terminal A is connected to the supply positive. Similarly, a two-quadrant drive will need only one leg of the bridge, where the upper device can be controlled for motoring mode and the lower device can be controlled for regeneration mode.

Converter Control of AC Machines

Although application of dc drives is quite common, disadvantages are that the machines are bulky and expensive, and the commutators and brushes require frequent maintenance. In fact, commutator sparking prevents machine application in an unclean environment, at high speed, and at high elevation. AC machines, particularly the cage-type induction motor, are favorable when compared with all the features of dc machines. Although converter system, control, and signal processing of ac drives is definitely complex, the evolution of ac drive technology in the past two decades has permitted more economical and higher performance ac drives. Consequently, ac drives are finding expanding applications, pushing dc drives towards obsolescence.

Voltage-Fed Inverter Induction Motor Drive

A simple and popular converter system for speed control of an induction motor is shown in Fig. 30.43. The front-end diode rectifier converts 60 Hz ac to dc, which is then filtered to remove the ripple. The dc voltage is then converted to variable-frequency, variable-voltage output for the machine through a PWM bridge inverter. Among a number of PWM techniques, the sinusoidal PWM is common, and it is illustrated in Fig. 30.44 for one phase only. The stator sinusoidal reference phase voltage signal is compared with a high-frequency carrier wave, and the comparator logic output controls switching of the upper and lower transistors in a phase leg. The phase voltage wave shown refers to the fictitious center tap of the filter capacitor. With the PWM technique, the fundamental voltage and frequency can be easily varied. The stator voltage wave contains high-frequency ripple, which is easily filtered by the machine leakage inductance. The voltage-to-frequency ratio is kept constant to provide constant airgap flux in the machine. The machine voltage-frequency relation, and the corresponding torque, stator current, and slip, are shown in Fig. 30.45. Up to the base or rated frequency ω_b, the machine can develop constant torque. Then, the field flux weakens as the frequency is increased at constant voltage. The speed of the machine can be controlled in a simple open-loop manner by controlling the frequency and maintaining the proportionality between the voltage and frequency. During acceleration, machine-developed torque should be limited so that the inverter current rating is not exceeded. By controlling the frequency, the operation can be extended in the field weakening region. If the supply frequency is controlled to be lower than the machine speed (equivalent frequency), the motor will act as a generator and the inverter will act as a rectifier, and energy from the motor will be pumped back to the dc link. The **dynamic brake** shown is nothing but a buck converter with resistive load that dissipates excess power to maintain the dc bus voltage constant. When

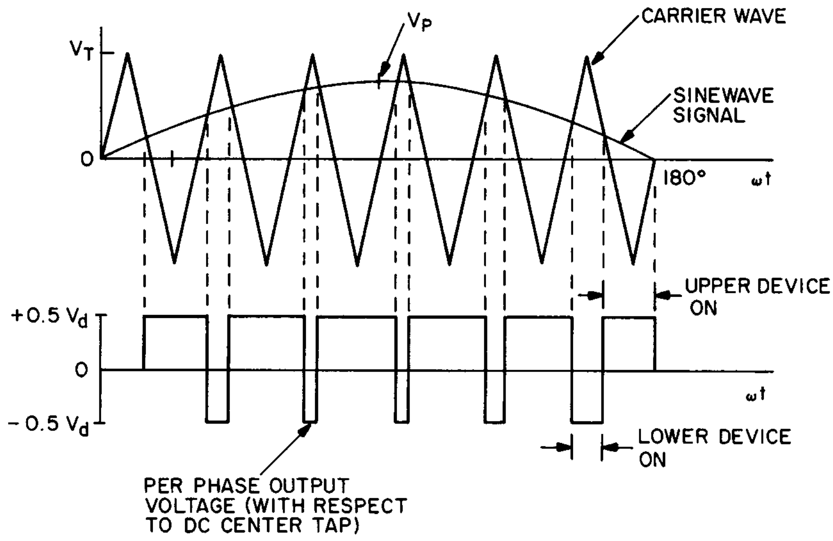


FIGURE 30.44 Sinusoidal pulse width modulation principle.

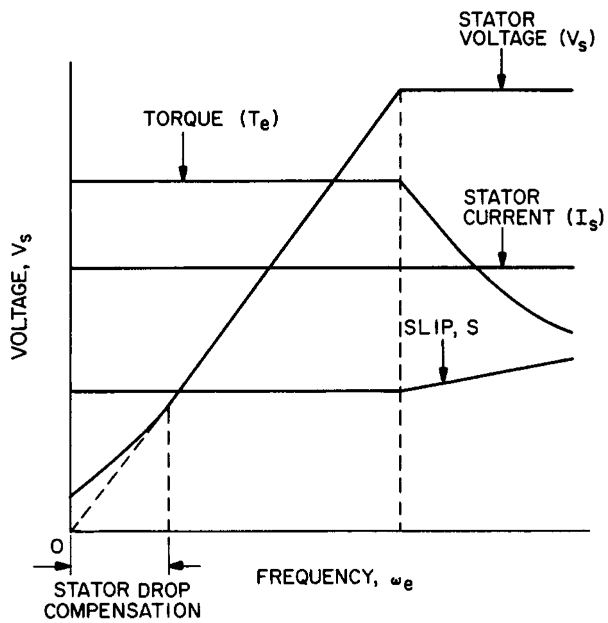


FIGURE 30.45 Voltage-frequency relation of an induction motor.

the motor speed is reduced to zero, the phase sequence of the inverter can be reversed for speed reversal. Therefore, the machine speed can be easily controlled in all four quadrants.

Current-Fed Inverter Induction Motor Drive

The speed of a machine can be controlled by a current-fed inverter as shown in Fig. 30.46. The front-end thyristor rectifier generates a variable dc current source in the dc link inductor. The dc current is then converted to six-step machine current wave through the inverter. The basic mode of operation of the inverter is the same as that of the rectifier, except that it is **force-commutated,** that is, the capacitors and series diodes help commutation of the thyristors. One advantage of the drive is that regenerative braking is easy because the

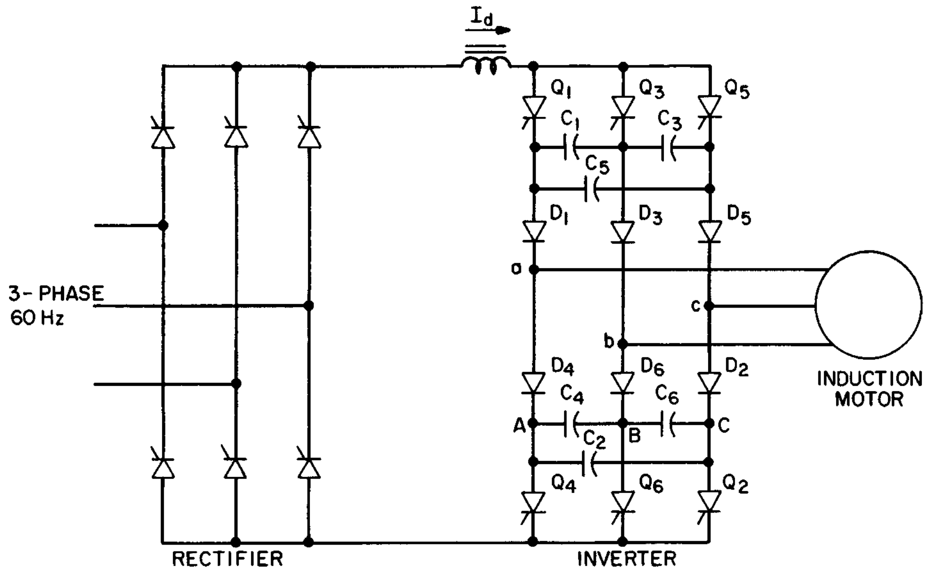


FIGURE 30.46 Force-commutated current-fed inverter control of an induction motor.

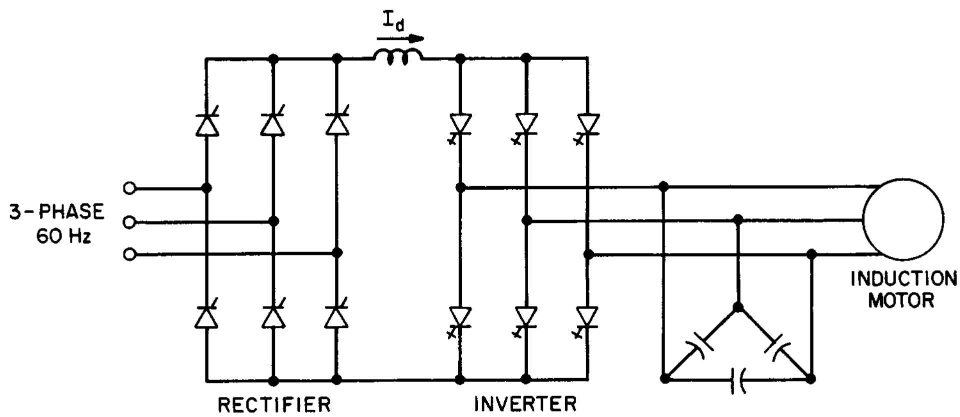


FIGURE 30.47 PWM current-fed inverter control of an induction motor.

rectifier and inverter can reverse their operation modes. Six-step machine current, however, causes large harmonic heating and torque pulsation, which may be quite harmful at low-speed operation. Another disadvantage is that the converter system cannot be controlled in open loop like a voltage-fed inverter.

Current-Fed PWM Inverter Induction Motor Drive

The force-commutated thyristor inverter in Fig. 30.46 can be replaced by a **self-commutating** gate turn-off (GTO) thyristor PWM inverter as shown in Fig. 30.47. The output capacitor bank shown has two functions: (1) it permits PWM switching of the GTO by diverting the load inductive current, and (2) it acts as a low-pass filter causing sinusoidal machine current. The second function improves machine efficiency and attenuates the irritating magnetic noise. Note that the fundamental machine current is controlled by the front-end rectifier, and the fixed PWM pattern is for controlling the harmonics only. The GTO is to be the reverse-blocking type. Such drives are popular in the multimegawatt power range. For lower power, an **insulated gate bipolar transistor (IGBT)** or transistor can be used with a series diode.

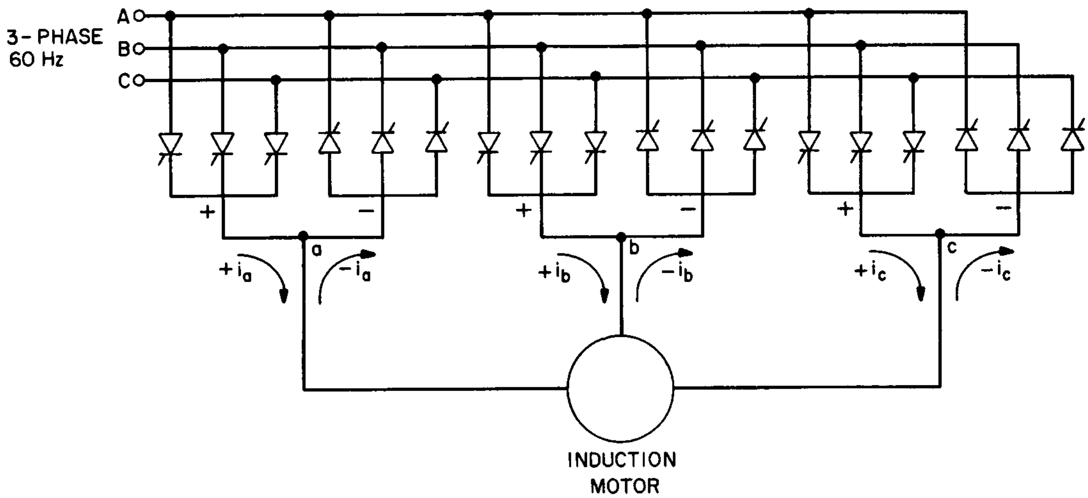


FIGURE 30.48 Cycloconverter control of an induction motor.

Cycloconverter Induction Motor Drive

A phase-controlled cycloconverter can be used for speed control of an ac machine (induction or synchronous type). Figure 30.48 shows a drive using a three-pulse half-wave or 18-thyristor cycloconverter. Each output phase group consists of positive and negative converter components which permit bidirectional current flow. The firing angle of each converter is sinusoidally modulated to generate the variable-frequency, variable-voltage output required for ac machine drive. Speed reversal and regenerative mode operation are easy. The cycloconverter can be operated in blocking or circulating current mode. In blocking mode, the positive or negative converter is enabled, depending on the polarity of the load current. In circulating current mode, the converter components are always enabled to permit circulating current through them. The circulating current reactor between the positive and negative converter prevents short circuits due to ripple voltage. The circulating current mode gives simple control and a higher range of output frequency with lower harmonic distortion.

Slip Power Recovery Drive of Induction Motor

In a cage-type induction motor, the rotor current at slip frequency reacting with the airgap flux develops the torque. The corresponding slip power is dissipated in the rotor resistance. In a wound rotor induction motor, the slip power can be controlled to control the torque and speed of a machine. Figure 30.49 shows a popular slip power-controlled drive, known as a static Kramer drive. The slip power is rectified to dc with a diode rectifier and is then pumped back to an ac line through a thyristor phase-controlled inverter. The method permits speed control in the subsynchronous speed range. It can be shown that the developed machine torque is proportional to the dc link current I_d and the voltage V_d varies directly with speed deviation from the synchronous speed. The current I_d is controlled by the firing angle of the inverter. Since V_d and V_I voltages balance at steady state, at synchronous speed the voltage V_d is zero and the firing angle is 90 degrees. The firing angle increases as the speed falls, and at 50% synchronous speed the firing angle is near 180 degrees. This is practically the lowest speed in static Kramer drive. The transformer steps down the inverter input voltage to get a 180-degree firing angle at lowest speed. The advantage of this drive is that the converter rating is low compared with the machine rating. Disadvantages are that the line power factor is low and the machine is expensive. For limited speed range applications, this drive has been popular.

Wound Field Synchronous Motor Drive

The speed of a wound field synchronous machine can be controlled by a current-fed converter scheme as shown in Fig. 30.46, except that the forced-commutation elements can be removed. The machine is operated at leading power factor by overexcitation so that the inverter can be load commutated. Because of the simplicity of converter topology and control, such a drive is popular in the multimegawatt range.

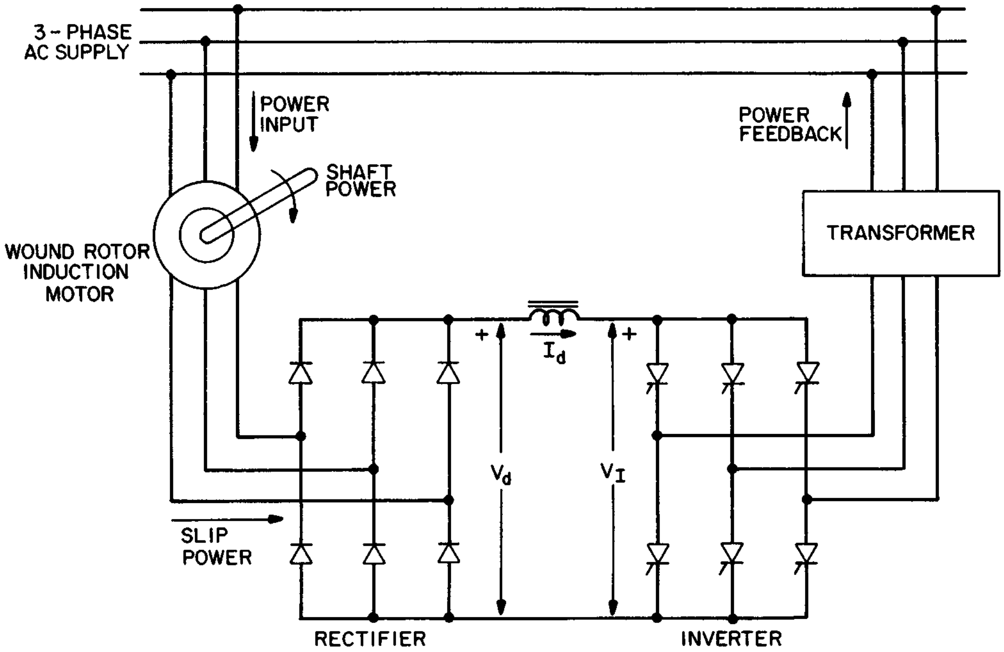


FIGURE 30.49 Slip power recovery control of a wound rotor induction motor.

Permanent Magnet Synchronous Motor Drive

Permanent magnet (PM) machine drives are quite popular in the low-power range. A PM machine can have sinusoidal or concentrated winding, giving the corresponding sinusoidal or trapezoidal induced stator voltage wave. Figure 30.50 shows the speed control system using a trapezoidal machine, and Fig. 30.51 explains the wave forms. The power MOSFET inverter supplies variable-frequency, variable-magnitude six-step current wave to the stator. The inverter is self-controlled, that is, the firing pulses are generated by the machine position sensor through a decoder. It can be shown that such a drive has the features of dc drive and is normally defined as *brushless dc drive*. The speed control loop generates the dc current command, which is then controlled by the **hysteresis-band** method to construct the six-step phase current waves in correct phase relation with the induced voltage waves as shown in Fig. 30.51. The drive can easily operate in four-quadrant mode.

Defining Terms

Dynamic brake: The braking operation of a machine by extracting electrical energy and then dissipating it in a resistor.

Forced-commutation: Switching off a power semiconductor device by external circuit transient.

Four-quadrant: A drive that can operate as a motor as well as a generator in both directions.

Hysteresis-band: A method of controlling current where the instantaneous current can vary within a band.

Insulated gate bipolar transistor (IGBT): A device that combines the features of a power transistor and MOSFET.

Regenerative braking: The braking operation of a machine by converting its mechanical energy into electrical form and then pumping it back to the source.

Self-commutation: Switching off a power semiconductor device by its gate or base drive.

Two-quadrant: A drive that can operate as a motor as well as a generator in one direction.

Related Topics

66.1 Generators • 66.2 Motors

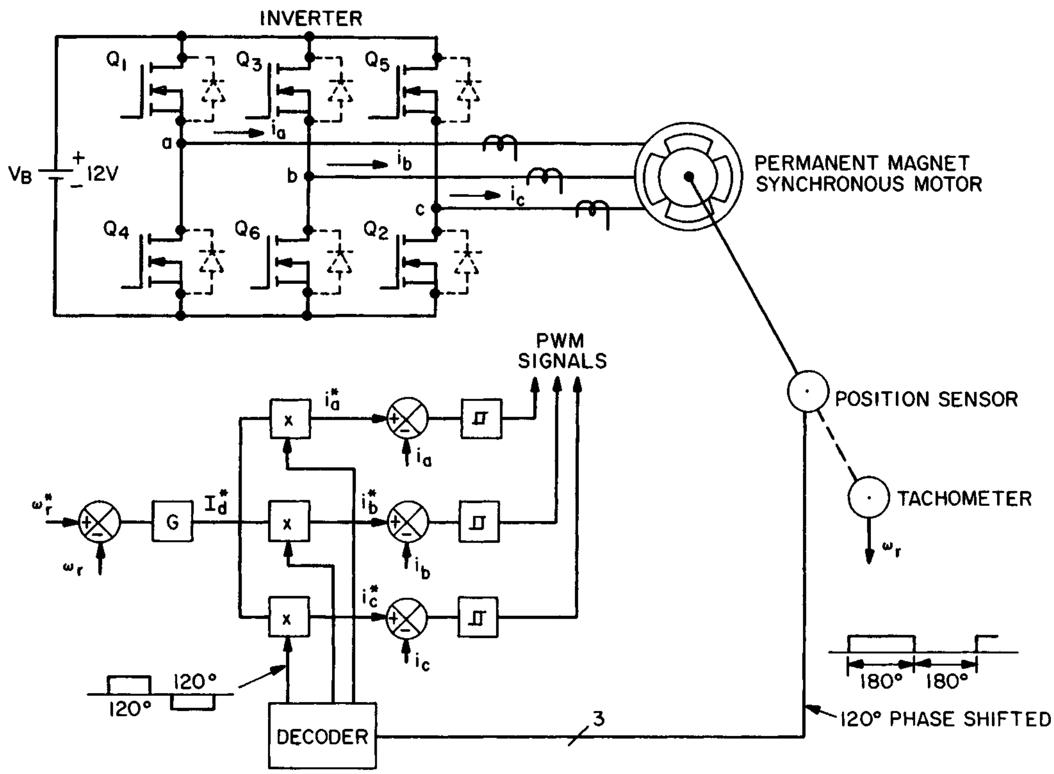


FIGURE 30.50 Permanent magnet synchronous motor control with PWM inverter.

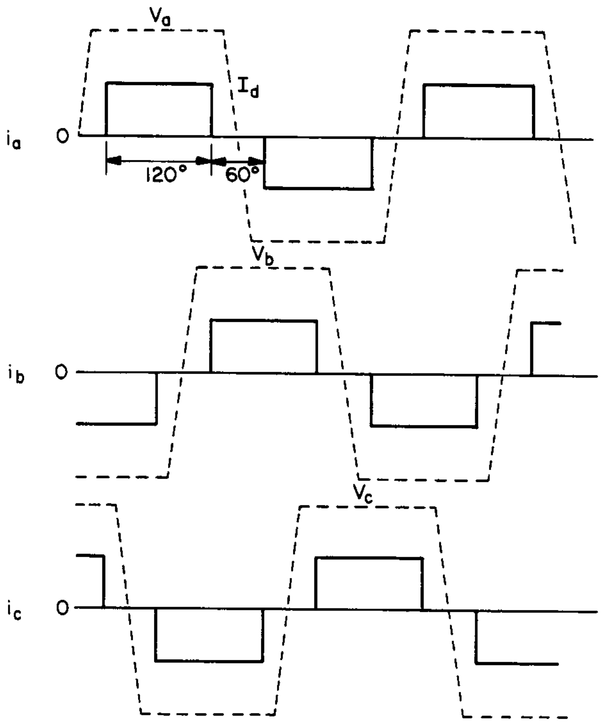


FIGURE 30.51 Phase voltage and current waves in brushless dc drive.

References

B.K. Bose, *Power Electronics and AC Drives*, Englewood Cliffs, N.J.: Prentice-Hall, 1986.
B.K. Bose, "Adjustable speed AC drives—A technology status review," *Proc. IEEE*, vol. 70, pp. 116–135, Feb. 1982.
B.K. Bose, *Modern Power Electronics*, New York: IEEE Press, 1992.
J.M.D. Murphy and F.G. Turnbull, *Power Electronic Control of AC Motors*, New York: Pergamon Press, 1988.
P.C. Sen, *Thyristor DC Drives*, New York: John Wiley, 1981.