UNIT IV
POWER ELECTRONICS
TE (E & TC)

EIGHT

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Inverters

A device that converts dc power into ac power at desired output voltage and frequency is called an inverter. Some industrial applications of inverters are for adjustable-speed ac drives, induction heating, stand by air-craft power supplies, UPS (uninterruptible power supplies) for computers, hvdc transmission lines etc. Phase-controlled converters, when operated in the inverter mode, are called line-commutated inverters, Chapter 6. But line-commutated inverters require at the output terminals an existing ac supply which is used for their commutation. This means that line-commutated inverters can't function as isolated ac voltage sources or as variable frequency generators with dc power at the input. Therefore, voltage level, frequency and waveform on the ac side of line-commutated inverters cannot be changed. On the other hand, force commutated inverters provide an independent ac output voltage of adjustable voltage and adjustable frequency and have therefore much wider applications. In this chapter, force-commutated and load commutated inverters are described.

The dc power input to the inverter is obtained from an existing power supply network or from a rotating alternator through a rectifier or a battery, fuel cell, photovoltaic array or magneto hydrodynamic (MHD) generator. The configuration of ac to dc converter and dc to ac inverter is called a dc-link converter. The rectification is carried out by standard diodes or thyristor converter circuits discussed in Chapter 6. The inversion is performed by the methods discussed in this chapter.

Inverters can be broadly classified into two types: voltage source inverters and current source inverters. A voltage-fed inverter (VFI), or voltage-source inverter (VSI), is one in which the dc source has small or negligible impedance. In other words, a voltage source inverter has stiff dc voltage source at its input terminals. A current-fed inverter (CFI) or current-source inverter (CSI) is fed with adjustable current from a dc source of high impedance, i.e. from a stiff dc current source. In a CSI fed with stiff current source, output current waves are not affected by the load.

In VSI's using thyristors, some type of forced commutation is usually required. In case VSI's are made up of using GTOs, power transistors, power MOSFETS or IGBTs, self-commutation with base or gate drive signals is employed for their controlled turn-on and turn-off.

The object of this chapter is to describe the operating principles of both single-phase and three-phase inverters and to present their elementary analysis. As before, switching devices are assumed to possess ideal characteristics.
.1. SINGLE-PHASE VOLTAGE SOURCE INVERTERS: OPERATING PRINCIPLE

In this section, operating principle of single-phase voltage source inverters is discussed.

8.1.1. Single-phase Bridge Inverters

Single-phase bridge inverters are of two types, namely (i) single-phase half-bridge inverters and (ii) single-phase full-bridge inverters. Basic principles of operation of these two types are presented here.

Power circuit diagrams of the two configurations of single-phase bridge inverter, as stated above, are shown in Fig. 8.1 (a) for half-bridge inverter and in Fig. 8.2 (a) for full-bridge inverter. In these diagrams, the circuitry for turning-on or turning-off of the thyristors is not shown for simplicity. The gating signals for the thyristors and the resulting output voltage waveforms are shown in Figs. 8.1 (b) and 8.2 (b) for half-bridge and full-bridge inverters respectively. These voltage waveforms are drawn on the assumption that each thyristor conducts for the duration its gate pulse is present and is commutated as soon as this pulse is removed. In Figs. 8.1 (b) and 8.2 (b), \( g_1 - g_4 \) are gate signals applied respectively to thyristors T1-T4.

Fig. 8.1. Single-phase half-bridge inverter.

Single-phase half bridge inverter, as shown in Fig. 8.1 (a), consists of two SCRs, two diodes and three-wire supply. It is seen from Fig. 8.1 (b) that for \( 0 < t \leq T/2 \), thyristor T1
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conducts and the load is subjected to a voltage $V_e/2$ due to the upper voltage source $V_e/2$. At $t = T/2$, thyristor T1 is commutated and T2 is gated on. During the period $T/2 < t < T$, thyristor T2 conducts and the load is subjected to a voltage $(-V_e/2)$ due to the lower voltage source $V_e/2$. It is seen from Fig. 8.1 (b) that load voltage is an alternating voltage waveform of amplitude $V_e/2$ and of frequency $1/T$ Hz. Frequency of the inverter output voltage can be changed by controlling $T$.

The main drawback of half-bridge inverter is that it requires 3-wire dc supply. This difficulty can, however, be overcome by the use of a full-bridge inverter shown in Fig. 8.2 (a). It consists of four SCRs and four diodes. In this inverter, number of thyristors and diodes is twice of that in a half-bridge inverter. This, however, does not go against full inverter because the amplitude of output voltage as well as its output power is doubled in this inverter as compared to their values in the half-bridge inverter. This is evident from Figs. 8.1 (b) and 8.2 (b).

For full-bridge inverter, when T1, T2 conduct, load voltage is $V_e$ and when T3, T4 conduct load voltage is $-V_e$ as shown in Fig. 8.2 (b). Frequency of output voltage can be controlled by varying the periodic time $T$.

In Fig. 8.1 (a), thyristors T1, T2 are in series across the source; in Fig. 8.2 (a) thyristors T1, T4 or T3, T2 are also in series across the source. During inverter operation, it should be ensured that two SCRs in the same branch, such as T1, T2 in Fig. 8.1 (a), do not conduct simultaneously as this would lead to a direct short circuit of the source.

For a resistive load, two SCRs in Fig. 8.1 (a) and four SCRs in Fig. 8.2 (a) would suffice, because load current $i_0$ and load voltage $v_0$ would always be in phase with each other. This, however, is not the case when the load is other than resistive. For such types of loads, current $i_0$ will not be in phase with voltage $v_0$ and diodes connected in antiparallel with thyristors will allow the current to flow when the main thyristors are turned off. These diodes are called feedback diodes.
8.4. THREE PHASE BRIDGE INVERTERS

For providing adjustable-frequency power to industrial applications, three-phase inverters are more common than single-phase inverters. Three-phase inverters, like single-phase inverters, take their dc supply from a battery or more usually from a rectifier.

A basic three-phase inverter is a six-step bridge inverter. It uses a minimum of six thyristors. In inverter terminology, a step is defined as a change in the firing from one thyristor to the next thyristor in proper sequence. For one cycle of 360°, each step would be of 60° interval for a six-step inverter. This means that thyristors would be gated at regular intervals of 60° in proper sequence so that a 3-phase ac voltage is synthesized at the output terminals of a six-step inverter.

Fig. 8.19 (a) shows the power circuit of a three-phase bridge inverter using six thyristors and six diodes. As stated earlier, the transistor family of devices is now very widely used in inverter circuits. Presently, the use of IGBTs in single-phase and three-phase inverters is on the rise. The basic circuit configuration of inverter, however, remains unaltered as shown in Fig. 8.19 (b) for a three-phase bridge inverter using IGBTs in place of thyristors. A large capacitor connected at the input terminals tends to make the input dc voltage constant. This capacitor also suppresses the harmonics fed back to the source.

In Fig. 8.19 (a) inverter using six thyristors, commutation and snubber circuits are omitted for simplicity. It may be seen from Figs. 8.1 and 8.19 that a three-phase bridge inverter consists of three half-bridge inverters arranged side by side. The three-phase load is assumed to be star connected. In Fig. 8.19 (a), the thyristors are numbered in the sequence in which they are triggered to obtain voltages $v_{ab}$, $v_{bc}$, $v_{ca}$ at the output terminals $a$, $b$, $c$ of the inverter.

There are two possible patterns of gating the thyristors. In one pattern, each thyristor conducts for 180° and in the other, each thyristor conducts for 120°. But in both these patterns,

![Diagram](image)

Fig. 8.19. Three-phase bridge inverter using (a) thyristors (b) IGBTs.
gating signals are applied and removed at 60° intervals of the output voltage waveform. Therefore, both these modes require a six step bridge inverter. These modes of thyristor conduction are described in what follows:

8.4.1. Three-phase 180 Degree Mode VSI

In the three-phase inverter of Fig. 8.19, each SCR conducts for 180° of a cycle. Thyristor pair in each arm, i.e. T1, T4; T3, T6 and T5, T2 are turned on with a time interval of 180°.

Fig. 8.20. Voltage waveforms for 180° mode 3-phase VSI.
It means that T1 conducts for 180° and T4 for the next 180° of a cycle. Thyristors in the upper group, i.e. T1, T3, T5 conduct at an interval of 120°. It implies that if T1 is fired at \( \omega t = 0° \), then T3 must be fired at \( \omega t = 120° \) and T5 at \( \omega t = 240° \). Same is true for lower group of SCRs. On the basis of this firing scheme, a table is prepared as shown at the top of Fig. 8.20. In this table, first row shows that T1 from upper group conducts for 180°, T4 for the next 180° and then again T1 for 180° and so on. In the second row, T3 from the upper group is shown to start conducting 120° after T1 starts conducting. After T3 conduction for 180°, T6 conducts for the next 180° and again T3 for the next 180° and so on. Further, in the third row, T5 from the upper group starts conducting 120° after T3 or 240° after T1. After T5 conduction for 180°, T2 conducts for the next 180°, T5 for the next 180° and so on. In this manner, the pattern of firing the six SCRs is identified. This table shows that T5, T6, T1 should be gated for step I, T6, T1, T2 for step II, T1, T2, T3 for step III, T2, T3, T4 for step IV and so on. Thus the sequence of firing the thyristors is T1, T2, T3, T4, T5, T6; T1, T2,... It is seen from the table that in every step of 60° duration, only three SCRs are conducting—one from upper group and two from the lower group or two from the upper group and one from the lower group.

The circuit models for step I-IV are shown in Fig. 8.21. During step I, thyristors 5, 6, 1 are conducting. These are shown as closed switches and non-conducting SCRs 2, 3, 4 as open switches in Fig. 8.21 (a). The load terminals a and c are connected to the positive bus of dc source whereas terminal b is connected to the negative bus of dc source, Fig. 8.21 (a). The load voltage is \( v_{ab} = v_{cb} = V_s \) in magnitude. The magnitude of line to neutral voltage can be obtained as under:

During step I, \( 0 \leq \omega t < \frac{\pi}{3} \), Fig. 8.21 (a), thyristors conducting 5, 6, 1.

Current,

\[
i_1 = \frac{V_s}{Z + Z} = \frac{2}{3} \cdot \frac{V_s}{Z}
\]

The line to neutral voltages are

\[
v_{ao} = v_{co} = i_1 \frac{Z}{2} = \frac{V_s}{3}
\]

and

\[
v_{ob} = i_1 \frac{Z}{2} = \frac{2V_s}{3}.
\]

The above line to neutral voltages may be written as \( v_{ao} = v_{co} = \frac{V_s}{3} \) and \( v_{ob} = -\frac{2V_s}{3} \). These voltages are shown in Fig. 8.20 (a) during step I. For the next step II, the line to neutral voltages are as under:

During step II, \( \frac{\pi}{3} \leq \omega t < \frac{2\pi}{3} \), Fig. 8.21 (b), thyristor conducting 6, 1, 2.

Current,

\[
i_2 = \frac{2V_s}{3} \frac{Z}{Z}
\]

\[
. \quad v_{ao} = i_2 \frac{Z}{2} = \frac{2V_s}{3} \quad ; \quad v_{ob} = v_{cb} = i_2 \frac{Z}{2} = \frac{V_s}{3}
\]

or

\[
v_{ao} = \frac{2V_s}{3} \quad ; \quad v_{bo} = v_{co} = -\frac{V_s}{3}.
\]
These output voltages are plotted in Fig. 8.20 (a). In this manner, the variation of phase voltages $v_{ao}, v_{bo}, v_{co}$ as obtained in Fig. 8.21 up to step IV and similarly for other steps, is plotted in Fig. 8.20 (a). It is clear that for each cycle of output voltage of each phase, six steps are required and each step has a duration of 60°.

**Step I**

(a) $0-60°$; 5, 6, 1 closed.

**Step II**

(b) $60-120°$; 6, 1, 2 closed.

**Step III**

(c) $120-180°$; 1, 2, 3 closed.

**Step IV**

(d) $180-240°$; 2, 3, 4 closed.

Fig. 8.21. Equivalent circuits for a 3-phase six-step 180° mode inverter with a balanced star-connected load.
The line voltage $v_{ab} = v_{ca} + v_{bc}$ or $v_{ab} = v_{ca} - v_{bc}$ is obtained by reversing $v_{bc}$ and adding it to $v_{ca}$ as shown in Fig. 8.20 (b). Similarly, line voltages $v_{bc} = v_{bo} - v_{co}$ and $v_{ca} = v_{co} - v_{ao}$ are plotted in Fig. 8.20 (b).

The three rows in the top of Fig. 8.20 also indicate the pattern of gating signal waveforms. At $\omega t = \pi$, when $i_{g1}$ is removed, T1 is turned off and simultaneously $i_{g4}$ is applied to turn on T4. Similarly, at $\omega t = 2\pi/3$, when $i_{g6}$ is cut off, T6 is turned off and at the same instant $i_{g3}$ is applied to turn on T3. Same is true for other thyristors.

It is seen from Fig. 8.20 that phase voltages have six steps per cycle and line voltages have one positive pulse and one negative pulse (each of 120° duration) per cycle. The phase as well as line voltages are out of phase by 120°. The function of diodes D1 to D6 is to allow the flow of currents through them when the load is reactive in nature.

The three line output voltages can be described by the Fourier series as follows:

$$v_{ob} = \sum_{n=1,3,5} \frac{4V_s}{n\pi} \cos \frac{n\pi}{6} \sin n(\omega t + \pi/6) \quad ... (8.44)$$

$$v_{bc} = \sum_{n=1,3,5} \frac{4V_s}{n\pi} \cos \frac{n\pi}{6} \sin n(\omega t - \pi/2) \quad ... (8.45)$$

$$v_{ca} = \sum_{n=1,3,5} \frac{4V_s}{n\pi} \cos \frac{n\pi}{6} \sin \left(\omega t + \frac{5\pi}{6}\right) \quad ... (8.46)$$

For $n = 3, \cos \frac{3\pi}{6} = 0$. Thus, all triplen harmonics are absent from the line voltages as given by Eqs. (8.44) to (8.46).

The line voltage waveforms shown in Fig. 8.20 represent a balanced set of three-phase alternating voltages. During the six intervals, these voltages are well defined. Therefore, these voltages are independent of the nature of load circuit which may consist of any combination of resistance, inductance and capacitance and the load may be balanced or unbalanced, linear or nonlinear.

Fourier series expansion of line to neutral voltage $v_{ao}$ in Fig. 8.20 is given by

$$v_{ao} = \sum_{n=-k \pm 1}^{\infty} \frac{2V_s}{n\pi} \sin n\omega t \quad ... (8.47)$$

where

$k = 0, 1, 2, ...$

For a linear star-connected balanced load, phase or line currents can be obtained from Eq. (8.47). Expressions similar to Eq. (8.47) can be written for $v_{bc}$ and $v_{ca}$ by replacing $\omega t$ by $(\omega t - 120^\circ)$ and $(\omega t - 240^\circ)$ respectively.

In Fig. 8.21, load is assumed star connected and three phase and line voltages are obtained as shown in Fig. 8.20. For a delta connected load also, phase or line voltage waveforms $v_{ab}, v_{bc}, v_{ca}$ as shown in Fig. 8.20 would be obtained directly. Therefore, for a linear delta-connected load, phase and line currents can be obtained from Eqs. (8.44) to (8.46). From Eq. (8.44), rms value of nth component of line voltage is

$$V_{Ln} = \frac{4V_s}{\sqrt{2}} \cos \frac{n\pi}{6} \quad ... (8.48)$$
Rms value of fundamental line voltage,

$$V_{L1} = \frac{4V_s}{\sqrt{2}} \cdot \cos \frac{\pi}{6} = 0.7797 V_s \quad \text{(8.49)}$$

It is seen from line voltage waveform $v_{ab}$ in Fig. 8.20 (a) that line voltage is $V_s$ from 0° to 120°. Therefore, rms value of line voltage $V_L$ is

$$V_L = \left[ \frac{1}{\pi} \int_0^{2\pi/3} V_s^2 d(\omega t) \right]^{1/2} = \sqrt{\frac{2}{3}} V_s = 0.8165 V_s \quad \text{(8.50)}$$

Rms value of phase voltage $V_p$ is

$$V_p = \frac{V_L}{\sqrt{3}} = \frac{\sqrt{2}}{3} V_s = 0.4714 V_s \quad \text{(8.51)}$$

Rms value of fundamental phase voltage, from Eq. (8.47), is

$$V_{p1} = \frac{2V_s}{\sqrt{2} \pi} = 0.4502 V_s = \frac{V_{L1}}{\sqrt{3}} \quad \text{(8.52)}$$

### 8.4.2. Three-phase 120 Degree Mode VSI

The power circuit diagram of this inverter is the same as that shown in Fig. 8.19. For the 120-degree mode VSI, each thyristor conducts for 120° of a cycle. Like 180° mode, 120° mode inverter also requires six steps, each of 60° duration, for completing one cycle of the output ac voltage.

For this inverter too, a table giving the sequence of firing the six thyristors is prepared as shown in the top of Fig. 8.22. In this table, first rows shows that T1 conducts for 120° and for the next 60°, neither T1 nor T4 conducts. Now T4 is turned on at $\omega t = 180°$ and it further conducts for 120°, i.e., from $\omega t = 180°$ to $\omega t = 300°$. This means that for 60° interval from $\omega t = 120°$ to $\omega t = 180°$, series connected SCRs do not conduct. At $\omega t = 300°$, T4 is turned off, then 60° interval elapses before T1 is turned on again at $\omega t = 360°$. In the second row, T3 is turned on at $\omega t = 120°$ as in 180° mode inverter. Now T3 conducts for 120°, then 60° interval elapses during which neither T3 nor T6 conducts. At $\omega t = 300°$, T6 is turned on, it conducts for 120° and then 60° interval elapses after which T3 is turned on again. The third row is also completed similarly. This table shows that T6, T1 should be gated for step I; T1, T2 for step II; T2, T3 for step III and so on. The sequence of firing the six thyristors is the same as for the 180° mode inverter. During each step, only two thyristors conduct for this inverter— one from the upper group and one from the lower group; but in 180° mode inverter, three thyristors conduct in each step.

The circuit models for steps I-IV are shown in Fig. 8.23, where load is assumed to be resistive and star connected. During step I, thyristors 6, 1 are conducting and as such load terminal $a$ is connected to the positive bus of dc source whereas terminal $b$ is connected to negative bus of dc source, Fig. 8.23 (a). Load terminal $c$ is not connected to dc bus. The line to neutral voltages, from Fig. 8.23 (a) are

$$v_{ad} = \frac{V_s}{2}, \quad v_{ab} = \frac{V_s}{2}$$

or

$$v_{bd} = -\frac{V_s}{2}$$

and

$$v_{ca} = 0$$
These voltages are shown in Fig. 8.22 (a) during step I of 0° – 60°. For step II, thyristors 1, 2 conduct and load voltages are \( v_{ac} = V_s/2 \), \( v_{ca} = -V_s/2 \) and \( v_{bc} = 0 \), Fig. 8.23 (b); these voltages are plotted in Fig. 8.22 (a). This procedure is followed for obtaining load voltages for the remaining steps and these phase voltages are then plotted in Fig. 8.22 (a).

The line voltages

\[ v_{ab} = v_{ac} - v_{bc} \]
\[ v_{bc} = v_{bc} - v_{ca} \]
\[ v_{ca} = v_{ca} - v_{ab} \]

are also plotted in Fig. 8.19 (b).

It is seen from Fig. 8.22 that phase voltages have one positive pulse and one negative pulse (each of 120° duration) for one cycle of output alternating voltage. The line voltages, however, have six steps per cycle of output alternating voltage.

As stated before, the three rows in the top of Fig. 8.22 indicate the pattern of gating signal waveforms.

The merits and demerits of 120-degree mode inverter over 180-degree mode inverter are as follows:
(a) 0—60°; 6, 1 closed

(b) 60—120°; 1, 2 closed

(c) 120—180°; 2, 3 closed

(d) 180—240°; 3, 4 closed

Fig. 8.23. Equivalent circuits for a 3-phase six-step 120° mode inverter with balanced star-connected resistive load.

(i) In the 180° mode inverter, when gate signal $i_{g1}$ is cut-off to turn off T1 at $\omega t = 180^\circ$, gating signal $i_{g4}$ is simultaneously applied to turn on T4 in the same leg. In practice, a
commutation interval must exist between the removal of \(i_{e1}\) and application of \(i_{e4}\), because otherwise dc source would experience a direct short-circuit through SCRs T1 and T4 in the same leg.

This difficulty is overcome considerably in 120-degree mode inverter. In this inverter, there is a 60\(^\circ\) interval between the turning off of T1 and turning on of T4. During this 60\(^\circ\) interval, T1 can be commutated safely. In general, this angular interval of 60\(^\circ\) exists between the turning-off of one device and turning-on of the complementary device in the same leg. This 60\(^\circ\) period provides sufficient time for the outgoing thyristor to regain forward blocking capability.

(ii) In the 120\(^\circ\) mode inverter, the potentials of only two output terminals connected to the dc source are defined at any time of the cycle. The potential of the third terminal, pertaining to a particular leg in which neither device is conducting, is not well defined; its potential therefore depends on the nature of the load circuit. Thus, the analysis of the performance of this inverter is complicated for a general load circuit. For a balanced resistive load, the potential of all the three terminals is, however, well defined. This is the reason load is assumed resistive in Fig. 8.23. For a balanced delta-connected resistive load, the line voltages as shown in Fig. 8.22 (b) are obtained directly.

The Fourier analysis of phase voltage waveform \(v_{ao}\) of Fig. 8.22 (a) is

\[
v_{ao} = \sum_{n=1,3,5} \frac{2V_s}{n\pi} \cos \frac{n\pi}{6} \sin n(\omega t + \pi/6) \quad \ldots \text{(8.53)}
\]

Similarly,

\[
v_{bo} = \sum_{n=1,3,5,\ldots} \frac{2V_s}{n\pi} \cos \frac{n\pi}{6} \sin n(\omega t - \pi/2) \quad \ldots \text{(8.54)}
\]

and

\[
v_{co} = \sum_{n=1,3,5,\ldots} \frac{2V_s}{n\pi} \cos \frac{n\pi}{6} \sin n(\omega t + 5\pi/6) \quad \ldots \text{(8.55)}
\]

The Fourier analysis of line voltage waveform \(v_{ab}\) of Fig. 8.22 (b) is

\[
v_{ab} = \sum_{n=6k+1} \frac{3V_s}{n\pi} \sin n(\omega t + \pi/3) \quad \ldots \text{(8.56)}
\]

where

\[k = 0, 1, 2, 3\ldots\]

Similar expressions for \(v_{bc}\) and \(v_{ca}\) can also be written.

Rms value of fundamental phase voltage, from Eq. (8.53), is

\[
V_{p1} = \frac{2V_s}{\sqrt{2} \cdot \pi} \cos \frac{\pi}{6} = 0.3889 V_s \quad \ldots \text{(8.57)}
\]

Rms value of phase voltage,

\[
V_p = \left[\frac{1}{\pi} \int_0^{2\pi/3} \left(\frac{V_s}{2}\right)^2 \cdot d(\omega t)\right]^{1/2} = \sqrt{\frac{2}{3}} \cdot \frac{V_s}{\sqrt{6}} = 0.4082 V_s \quad \ldots \text{(8.58)}
\]

Rms value of fundamental line voltage, from Eq. (8.56), is

\[
V_{L1} = \frac{3V_s}{\sqrt{2} \cdot \pi} = 0.6752 V_s = \sqrt{3} V_{p1} \quad \ldots \text{(8.59)}
\]
Rms value of line voltage,
\[ V_L = \sqrt{3} \cdot V_p = \frac{V_s}{\sqrt{3}} = 0.7071 \cdot V_s \] \hspace{1cm} \text{(8.60)}

**Example 8.8.** A three-phase bridge inverter delivers power to a resistive load from a 450 V dc source. For a star-connected load of 10 Ω per phase, determine for both (a) 180° mode and (b) 120° mode,
(i) rms value of load current
(ii) rms value of thyristor current
(iii) load power.

**Solution.** For a resistive load, the waveform of load current is the same as that of the applied voltage. In view of this, waveforms of phase-load current and thyristor current are as shown in Fig. 8.24 (a) for 180° mode operation and in Fig. 8.24 (b) for 120° mode operation.

![Waveform Diagrams](image)

(a) \textbf{180° mode} : Upper waveform of Fig. 8.24 (a) shows that rms value of per-phase load current \( I_{or} \) is given by
\[
I_{or} = \left[ \frac{1}{\pi} \left( \frac{V_s}{3R} \right)^2 \frac{\pi}{3} + \left( \frac{2V_s}{3R} \right)^2 \frac{\pi}{3} + \left( \frac{V_s}{3R} \right)^2 \frac{\pi}{3} \right]^{1/2}
\]
\[
= \left[ \frac{450}{3 \times 10} \right]^2 \frac{2}{3} + \left( \frac{2 \times 450}{3 \times 10} \right)^2 \frac{1}{3} \right]^{1/2} = \sqrt{350} = 18.708 \text{ A}
\]

Rms value of thyristor current is
\[
I_{T1} = \left[ \frac{1}{2\pi} \left( \frac{450}{3 \times 10} \right)^2 \frac{2\pi}{3} + \left( \frac{2 \times 450}{3 \times 10} \right)^2 \frac{\pi}{3} \right]^{1/2}
\]
\[
= \sqrt{175} = 13.229 \text{ A}
\]

Power delivered to load
\[ = 3 I_{or}^2 R = 3 \left( \sqrt{350} \right)^2 \times 10 = 10.5 \text{ kW} \]

(b) \textbf{120° mode} : Upper waveform in Fig. 8.24 (b) gives rms value of per-phase load current \( I_{or} \) as under:
\[
I_{or} = \left[ \frac{1}{\pi} \left( \frac{450}{2 \times 10} \right)^2 \frac{2\pi}{3} \right]^{1/2} = \sqrt{337.5} = 18.371 \text{ A}
\]
8.7. REDUCTION OF HARMONICS IN THE INVERTER OUTPUT VOLTAGE

There are several industrial applications which may allow a harmonic content of 5% of its fundamental component of input voltage when inverters are used. Actually, the inverter output voltage may have harmonic content much higher than 5% of its fundamental component. In order to bring this harmonic content to a reasonable limit of 5%, one method is to insert filters between the load and inverter. If the inverter output voltage contains high-frequency harmonics, these can be reduced by a low-size filter. For the attenuation of low-frequency harmonics, however, the size of filter components increases. This makes the filter circuit costly, bulky and weighty and in addition, the transient response of the system becomes sluggish. This shows that lower order harmonics from the inverter output voltage should be reduced by some means other than the filter. Subsequent to this, high frequency component from this voltage can easily be attenuated by a low-size, low-cost filter. The object of this section is to study these methods of reducing low-order harmonics from the output voltage of an inverter.

8.7.1. Harmonic Reduction by PWM

It has already been discussed that when there are several pulses per half cycle, lower-order harmonics are eliminated. Fig. 8.34 illustrates output voltage waveform that can be obtained from a single-phase full-bridge inverter. This waveform can also be obtained from a single-phase half-bridge inverter, but then the amplitude of voltage wave would be $V_s/2$. The waveform of Fig. 8.34 needs ten commutations per cycle ($= 360^\circ$) instead of two in an unmodulated wave. The voltage waveform of Fig. 8.34 is symmetrical about $\pi$ as well as $\pi/2$.

\[
\begin{align*}
A_n &= \frac{4}{\pi} V_s \left[ \int_0^{\alpha_1} \sin n\omega t \cdot d(\omega t) - \int_{\alpha_1}^{\alpha_2} \sin n\omega t \cdot d(\omega t) + \int_{\alpha_2}^{\pi/2} \sin n\omega t \cdot d(\omega t) \right] \\
&= \frac{4V_s}{\pi} \left[ 1 - 2 \cos n\alpha_1 + 2 \cos n\alpha_2 \right]
\end{align*}
\]  

As this voltage waveform has quarter-wave symmetry, $B_n = 0$.

\[
\begin{align*}
A_n &= \frac{4}{\pi} V_s \left[ \int_0^{\alpha_1} \sin n\omega t \cdot d(\omega t) - \int_{\alpha_1}^{\alpha_2} \sin n\omega t \cdot d(\omega t) + \int_{\alpha_2}^{\pi/2} \sin n\omega t \cdot d(\omega t) \right] \\
&= \frac{4V_s}{\pi} \left[ 1 - 2 \cos n\alpha_1 + 2 \cos n\alpha_2 \right]
\end{align*}
\]
If third and fifth harmonics are to be eliminated, then from Eq. (8.76),

\[ A_3 = \frac{4V_s}{\pi} \left[ \frac{1 - 2 \cos 3 \alpha_1 + 2 \cos 3 \alpha_2}{3} \right] = 0 \]

and

\[ A_5 = \frac{4V_s}{\pi} \left[ \frac{1 - 2 \cos 5\alpha_1 + 2 \cos 5\alpha_2}{5} \right] = 0 \]

or

\[ 1 - 2 \cos 3\alpha_1 + 2 \cos 3\alpha_2 = 0 \]

and

\[ 1 - 2 \cos 5\alpha_1 + 2 \cos 5\alpha_2 = 0 \]

The above two simultaneous equations can be solved numerically to calculate \( \alpha_1 \) and \( \alpha_2 \) under the condition that \( 0 < \alpha_1 < 90^\circ \) and \( \alpha_1 < \alpha_2 < 90^\circ \). This gives \( \alpha_1 = 23.62^\circ \) and \( \alpha_2 = 33.304^\circ \).

With these values of \( \alpha_1, \alpha_2 \), the amplitudes of 7th, 9th and 11th harmonics, from Eq. (8.76), are as under:

\[ A_7 = \frac{4V_s}{7\pi} \left[ 1 - 2 \cos 7x 23.62 + 2 \cos 7x 33.304 \right] = 0.31555 \ V_s \]

\[ A_9 = \frac{4V_s}{9\pi} \left[ 1 - 2 \cos 9x 23.62 + 2 \cos 9x 33.304 \right] = 0.5202 \ V_s \]

and

\[ V_{11} = \frac{4V_s}{11\pi} \left[ 1 - 2 \cos 11x 23.62 + 2 \cos 11x 33.304 \right] = 0.3867 \ V_s \]

The amplitude of the fundamental component for these values of \( \alpha_1 \) and \( \alpha_2 \) is

\[ A_1 = \frac{4V_s}{\pi} \left[ 1 - 2 \cos 23.62 + 2 \cos 33.304 \right] = 1.0684 \ V_s \]

The amplitude of the fundamental component of unmodulated output voltage wave is

\[ A_{1,n} = \frac{4V_s}{\pi} = 1.27324 \ V_s \]

In terms of the fundamental component of unmodulated voltage wave, the amplitude of 7th, 9th and 11th harmonics are respectively 24.78% \( (= 0.31555 \times 100/1.27324) \), 40.86% and 30.37% but third and fifth harmonics are eliminated from the inverter output voltage wave. The amplitude of the fundamental voltage is 83.91% or 0.6391 times the amplitude of fundamental component of unmodulated voltage wave. Thus, with this method of harmonic reduction inverter is derated by (100-83.91) 16.09%. Another disadvantage of this method is that there are additional eight commutations per cycle and this leads to more switching losses in the thyristors.

### 8.7.2. Harmonic Reduction by Transformer Connections

Output voltage from two or more inverters can be combined by means of transformers to get a net output voltage with reduced harmonic content. The essential condition of this scheme is that the output voltage waveforms from the inverters must be similar but phase-shifted from each other. Fig. 8.35 (a) illustrates two transformers in series. Their output voltages, \( v_{01} \) from inverter 1 and \( v_{02} \) from inverter 2, are shown in Fig. 8.34 (b). Here \( v_{02} \) waveform is taken to have a phase shift of \( \pi/3 \) radians with respect to \( v_{01} \) waveform as shown. The resultant output voltage \( v_0 \) is obtained by adding the vertical ordinates of \( v_{01} \) and \( v_{02} \). It is seen that \( v_0 \) has an amplitude of \( 2V_s \) from \( \frac{\pi}{3} \) to \( \frac{4\pi}{3} \) to \( 2\pi \) and so on. Note that shape of the output voltage wave \( v_0 \) is a quasi-square wave.
Fig. 8.35. (a) Harmonic reduction by transformer connections (b) Elimination of third and other triplen harmonics.

The Fourier analysis of waveforms \( v_{01} \) and \( v_{02} \) gives

\[
v_{01} = \frac{4V_s}{\pi} \left[ \sin \omega t + \frac{1}{3} \sin 3\omega t + \frac{1}{5} \sin 5\omega t + \frac{1}{7} \sin 7\omega t + \ldots \right]
\]

\[
v_{02} = \frac{4V_s}{\pi} \left[ \sin \left( \omega t - \frac{\pi}{3} \right) + \frac{1}{3} \sin 3 \left( \omega t - \frac{\pi}{3} \right) + \frac{1}{5} \sin 5 \left( \omega t - \frac{\pi}{3} \right) + \frac{1}{7} \sin 7 \left( \omega t - \frac{\pi}{3} \right) + \ldots \right]
\]

The resultant voltage \( v_0 \) is

\[
v_0 = v_{01} + v_{02} = \frac{4V_s}{\pi} \sqrt{3} \left[ \sin \left( \omega t - \frac{\pi}{6} \right) + \frac{1}{5} \sin \left( 5\omega t + \frac{\pi}{6} \right) + \frac{1}{7} \sin \left( 7\omega t - \frac{\pi}{6} \right) + \ldots \right] \quad \ldots (8.77)
\]

The expression for resultant voltage \( v_0 \) as given above can be obtained from \( v_{01} \) and \( v_{02} \) analytically or graphically. The summation by graphical method is carried out as under:

An examination of the expressions for \( v_{01} \) and \( v_{02} \) reveals that for the fundamental frequency; \( V_{02} \) lags \( V_{01} \) by 60°, this is shown in Fig. 8.36 (a). The resultant of \( V_{01} \) and \( V_{02} \) must be \( \sqrt{3} \) times \( V_{01} \) (or \( V_{02} \)) and at the same time, the resultant lags \( V_{01} \) by 30°. Net value of fundamental frequency voltage must, therefore, be associated with \( \sqrt{3} \sin (\omega t - \pi/6) \). For

Fig. 8.36. Pertaining to the summation of first, third, fifth and seventh harmonic voltages.
third harmonic, \( V_{o2} \) lags \( V_{o1} \) by 180°, Fig. 8.36 (b), their resultant is therefore zero. For fifth harmonic, \( V_{o2} \) lags \( V_{o1} \) by 300° or \( V_{o2} \) leads \( V_{o1} \) by 60°, see Fig. 8.36 (c); its resultant is \( \sqrt{3} \) times \( V_{o1} \) or \( V_{o2} \) and it leads \( V_{o1} \) by 30°. Thus, the resultant of fifth harmonic voltage must be associated with \( \sqrt{3} \sin (\omega t + \pi/6) \). Similarly, resultant of seventh harmonic voltage must be associated with \( \sqrt{3} \sin (\omega t - \pi/6) \), Fig. 8.36 (d).

It is seen from Eq. (8.77) that third and other triplen harmonics are eliminated from net output voltage wave. The amplitude of fundamental component of \( v_o \) is

\[
V_{o1m} = \frac{4V_s}{\pi} \sqrt{3}
\]

In case output voltages \( v_{o1}, v_{o2} \) from inverters 1 and 2 has no phase shift, then amplitude of the fundamental voltage wave is \( 8V_s/\pi \). This shows that with phase shift, the amplitude of the fundamental voltage is \( \frac{4V_s}{\pi} \cdot \sqrt{3} \times \frac{\pi}{8V_s} = \frac{\sqrt{3}}{2} \) times the amplitude of fundamental voltage with no phase shift. With this method of harmonic reduction, there is thus a derating of

\[
\left( \frac{8V_s}{\pi} I - \frac{4V_s}{\pi} \cdot \sqrt{3} I \right) 100 = \left( 1 - \frac{\sqrt{3}}{2} \right) 100 = 13.4\%
\]

in their net output power so far as fundamental component is concerned. The degree of derating with this method is, however, less than that obtained in PWM harmonic reduction method.

The disadvantage of this method of harmonic reduction is the need for more number of inverters and transformers of similar ratings.

8.7.3. Harmonic Reduction by Stepped-wave Inverters

In this method, pulses of different widths and heights are superimposed to produce a resultant stepped wave with reduced harmonic content. Fig. 8.37 illustrates two stepped-wave inverters fed from a common dc supply. The two transformers used have different turns ratio from primary to secondary. In this figure, the turns ratio from primary to secondary is assumed three for transformer 1 and unity for transformer 2.
The inverter I is so gated that its output voltage is \( v_{o1} \) as shown in Fig. 8.38 (a). During the first-half cycle, output voltage level is either zero or positive. During second half cycle (not shown in the figure), the output voltage would be either zero or negative. This output voltage waveform is given the name two-level modulation.

For inverter II, the triggering is so arranged as to give output voltage \( v_{o2} \) as shown in Fig. 8.38 (b). It is seen from \( v_{o2} \) waveform that the level of output voltage is positive, negative or zero during the first half cycle, this inverter has therefore three-level modulation. The resultant output voltage from a series combination of inverters I and II is obtained by superimposing the waveforms of Figs. 8.38 (a) and (b). This summation depicted in Fig. 8.38 (c) shows that the amplitude of output voltage is \( 4V_s \) and waveform has four steps. Fourier analysis of Fig. 8.38 (c) would give harmonics whose amplitudes would depend upon the values of \( d_1, d_2, d_3, d_4 \) and amplitude of \( v_o \). By a proper choice of these parameters; third, fifth and seventh harmonics can be eliminated or attenuated considerably and the fundamental component optimised.

Note that it is the three-level modulation of second inverter that helps in achieving the required wave-stepping of the resultant output voltage waveform.

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3.4 PARALLEL INVERTER (WITH PURELY RESITIVE LOAD):

![Diagram of a parallel inverter with thyristors and transformer.]

Fig. 5.5.1 (a) Circuit Diagram.

The parallel inverter is sometimes also called as the centre tapped inverter because this configuration needs a centre tapped transformer on the output side.

The thyristors $S_1$ and $S_2$ are switched alternately to connect the input dc source $V$ in alternative senses across the two halves of the transformer primary. This induces a square wave voltage across the load in the transformer secondary.

$C$ is the commutation capacitor. The voltage on the capacitor is used to turn off a conducting SCR, by turning on the nonconducting SCR.

When $SCR_1$ is turned on, the dc source voltage appears across the left half of the primary OA. The primary current flows from O to A. Due to the transformer action the voltage between AB is $2V$. Hence the capacitor is
charged to a voltage of 2V. The load voltage is positive, so is the load current. (Fig. 5.5.1 (b)).

![Mode I equivalent circuit.]

**Mode II:**

The firing of SCR₂ turns off SCR₁ by the principle of parallel capacitor commutation. (The capacitor voltage is applied across SCR₁ directly to reverse bias it). The input dc voltage now gets connected across winding OB. The primary current flows from O to B, through SCR₂ as shown in Fig. 5.5.1 (c). The load voltage changes its polarity, and the direction of load current is reversed.

![Mode II equivalent circuit.]

**Fig. 5.5.1 (c) MODE II equivalent circuit.**

The square output waveform is thus obtained across the load.
Fig. 5.5.1 (d) Waveforms for parallel inverter with resistive load.

P.J. Shah
CHAPTER 8

✓ SWITCH-MODE dc±ac INVERTERS: dc ↔ SINUSOIDAL ac

P.J. Shah

8-1 INTRODUCTION

Switch-mode dc-to-ac inverters are used in ac motor drives and uninterruptible ac power supplies where the objective is to produce a sinusoidal ac output whose magnitude and frequency can both be controlled. As an example, consider an ac motor drive, shown in Fig. 8-1 in a block diagram form. The dc voltage is obtained by rectifying and filtering the line voltage, most often by the diode rectifier circuits discussed in Chapter 5. In an ac motor load, as will be discussed in Chapters 14 and 15, the voltage at its terminals is desired to be sinusoidal and adjustable in its magnitude and frequency. This is accomplished by means of the switch-mode dc-to-ac inverter of Fig. 8-1, which accepts a dc voltage as the input and produces the desired ac voltage input.

To be precise, the switch-mode inverter in Fig. 8-1 is a converter through which the power flow is reversible. However, most of the time the power flow is from the dc side to the motor on the ac side, requiring an inverter mode of operation. Therefore, these switch-mode converters are often referred to as switch-mode inverters.

To slow down the ac motor in Fig. 8-1, the kinetic energy associated with the inertia of the motor and its load is recovered and the ac motor acts as a generator. During the so-called braking of the motor, the power flows from the ac side to the dc side of the switch-mode converter and it operates in a rectifier mode. The energy recovered during the braking of the ac motor can be dissipated in a resistor, which can be switched in

![Figure 8-1 Switch-mode inverter in ac motor drive.](image-url)
parallel with the dc bus capacitor for this purpose in Fig. 8-1. However, in applications where this braking is performed frequently, a better alternative is regenerative braking where the energy recovered from the motor load inertia is fed back to the utility grid, as shown in the system of Fig. 8-2. This requires that the converter connecting the drive to the utility grid be a two-quadrant converter with a reversible dc current, which can operate as a rectifier during the motoring mode of the ac motor and as an inverter during the braking of the motor. Such a reversible-current two-quadrant converter can be realized by two back-to-back connected line-frequency thyristor converters of the type discussed in Chapter 6 or by means of a switch-mode converter as shown in Fig. 8-2. There are other reasons for using such a switch-mode rectifier (called a rectifier because, most of the time, the power flows from the ac line input to the dc bus) to interface the drive with the utility system. A detailed discussion of switch-mode rectifiers is deferred to Chapter 18, which deals with issues regarding the interfacing of power electronics equipment with the utility grid.

In this chapter, we will discuss inverters with single-phase and three-phase ac outputs. The input to switch-mode inverters will be assumed to be a dc voltage source, as was assumed in the block diagrams of Fig. 8-1 and 8-2. Such inverters are referred to as voltage source inverters (VSIs). The other types of inverters, now used only for very high power ac motor drives, are the current source inverters (CSI), where the dc input to the inverter is a dc current source. Because of their limited applications, the CSIs are not discussed in this chapter, and their discussion is deferred to ac motor drives Chapters 14 and 15.

The VSIs can be further divided into the following three general categories:

1. Pulse-width-modulated inverters. In these inverters, the input dc voltage is essentially constant in magnitude, such as in the circuit of Fig. 8-1, where a diode rectifier is used to rectify the line voltage. Therefore, the inverter must control the magnitude and the frequency of the ac output voltages. This is achieved by PWM of the inverter switches and hence such inverters are called PWM inverters. There are various schemes to pulse-width modulate the inverter switches in order to shape the output ac voltages to be as close to a sine wave as possible. Out of these various PWM schemes, a scheme called the sinusoidal PWM will be discussed in detail, and some of the other PWM techniques will be described in a separate section at the end of this chapter.

2. Square-wave inverters. In these inverters, the input dc voltage is controlled in order to control the magnitude of the output ac voltage, and therefore the inverter has to control only the frequency of the output voltage. The output ac voltage has a waveform similar to a square wave, and hence these inverters are called square-wave inverters.

![Figure 8-2 Switch-mode converters for motoring and regenerative braking in ac motor drive.](attachment://image.png)
3. Single-phase inverters with voltage cancellation. In case of inverters with single-phase output, it is possible to control the magnitude and the frequency of the inverter output voltage, even though the input to the inverter is a constant dc voltage and the inverter switches are not pulse-width modulated (and hence the output voltage waveshape is like a square wave). Therefore, these inverters combine the characteristics of the previous two inverters. It should be noted that the voltage cancellation technique works only with single-phase inverters and not with three-phase inverters.

8-2 BASIC CONCEPTS OF SWITCH-MODE INVERTERS

In this section, we will consider the requirements on the switch-mode inverters. For simplicity, let us consider a single-phase inverter, which is shown in block diagram form in Fig. 8-3a, where the output voltage of the inverter is filtered so that \( v_o \) can be assumed to be sinusoidal. Since the inverter supplies an inductive load such as an ac motor, \( i_o \) will lag \( v_o \), as shown in Fig. 8-3b. The output waveforms of Fig. 8-3b show that during interval 1, \( v_o \) and \( i_o \) are both positive, whereas during interval 3, \( v_o \) and \( i_o \) are both negative. Therefore, during intervals 1 and 3, the instantaneous power flow \( p_o (=v_o i_o) \) is from the dc side to the ac side, corresponding to an inverter mode of operation. In contrast, \( v_o \) and \( i_o \) are of opposite signs during intervals 2 and 4, and therefore \( p_o \) flows from the ac side to the dc side of the inverter, corresponding to a rectifier mode of operation. Therefore, the switch-mode inverter of Fig. 8-3a must be capable of operating in all four quadrants of the \( i_o-v_o \) plane, as shown in Fig. 8-3c during each cycle of the ac

![Diagram](image-url)
output. Such a four-quadrant inverter was first introduced in Chapter 7, where it was shown that in a full-bridge converter of Fig. 7-27, \( i_o \) is reversible and \( v_o \) can be of either polarity independent of the direction of \( i_o \). Therefore, the full-bridge converter of Fig. 7-27 meets the switch-mode inverter requirements. Only one of the two legs of the full-bridge converter, for example leg A, is shown in Fig. 8-4. All the dc-to-ac inverter topologies described in this chapter are derived from the one-leg converter of Fig. 8-4. For ease of explanation, it will be assumed that in the inverter of Fig. 8-4, the midpoint "o" of the dc input voltage is available, although in most inverters it is not needed and also not available.

To understand the dc-to-ac inverter characteristics of the one-leg inverter of Fig. 8-4, we will first assume that the input dc voltage \( V_d \) is constant and that the inverter switches are pulse-width modulated to shape and control the output voltage. Later on, it will be shown that the square-wave switching is a special case of the PWM switching scheme.

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UNIT V
POWER ELECTRONICS
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CHAPTER 11

Ac Voltage Controllers

The learning objectives of this chapter are as follows:

- To understand the operation and characteristics of ac voltage controllers
- To understand the operation of matrix converters
- To learn the types of ac voltage controllers
- To understand the performance parameters of ac voltage controllers
- To learn the techniques for analysis and design of ac voltage controllers
- To learn the techniques for simulating controlled rectifiers by using SPICE
- To study effects of load inductance on the load current

11.1 INTRODUCTION

If a thyristor switch is connected between ac supply and load, the power flow can be controlled by varying the rms value of ac voltage applied to the load; and this type of power circuit is known as an ac voltage controller. The most common applications of ac voltage controllers are: industrial heating, on-load transformer connection changing, light controls, speed control of polyphase induction motors, and ac magnet controls. For power transfer, two types of control are normally used:

1. On-off control
2. Phase-angle control

In on-off control, thyristor switches connect the load to the ac source for a few cycles of input voltage and then disconnect it for another few cycles. In phase control, thyristor switches connect the load to the ac source for a portion of each cycle of input voltage.

The ac voltage controllers can be classified into two types: (1) single-phase controllers and (2) three-phase controllers, with each type subdivided into (a) unidirectional or half-wave control and (b) bidirectional or full-wave control. There are various configurations of three-phase controllers depending on the connections of thyristor switches.
Because the input voltage is ac, thyristors are line commutated; and phase-control thyristors, which are relatively inexpensive and slower than fast-switching thyristors, are normally used. For applications up to 400 Hz, if TRIACs are available to meet the voltage and current ratings of a particular application, TRIACs are more commonly used.

Due to line or natural commutation, there is no need of extra commutation circuitry and the circuits for ac voltage controllers are very simple. Due to the nature of output waveforms, the analysis for the derivations of explicit expressions for the performance parameters of circuits is not simple, especially for phase-angle-controlled converters with \( RL \) loads. For the sake of simplicity, resistive loads are considered in this chapter to compare the performances of various configurations. However, the practical loads are of the \( RL \) type and should be considered in the design and analysis of ac voltage controllers.

### 11.2 PRINCIPLE OF ON-OFF CONTROL

The principle of on-off control can be explained with a single-phase full-wave controller, as shown in Figure 11.1a. The thyristor switch connects the ac supply to load for a time \( t_n \); the switch is turned off by a gate pulse inhibiting for time \( t_0 \). The on-time \( t_n \) usually consists of an integral number of cycles. The thyristors are turned on at the

![Circuit Diagram](image)

(a) Circuit

![Waveforms](image)

(b) Waveforms

---

**Power factor, PF**

\[
PF = \sqrt{k}
\]

(c) Power factor

![Graph](image)

**FIGURE 11.1**

On-off control.
zero-voltage crossings of ac input voltage. The gate pulses for thyristors $T_1$ and $T_2$ and the waveforms for input and output voltages are shown in Figure 11.1b.

This type of control is applied in applications that have a high mechanical inertia and high thermal time constant (e.g., industrial heating and speed control of motors). Due to zero-voltage and zero-current switching of thyristors, the harmonics generated by switching actions are reduced.

For a sinusoidal input voltage, $v_i = V_m \sin \omega t = \sqrt{2} V_s \sin \omega t$. If the input voltage is connected to load for $n$ cycles and is disconnected for $m$ cycles, the rms output (or load) voltage can be found from

$$V_o = \sqrt{\frac{n}{2\pi(n + m)} \int_0^{2\pi} 2V_s^2 \sin^2 \omega t \, d(\omega t)}$$

$$= V_s \sqrt{\frac{n}{m + n}} = V_s \sqrt{k}$$  \hspace{1cm} (11.1)

where $k = n/(m + n)$ and $k$ is called the duty cycle. $V_s$ is the rms phase voltage. The circuit configurations for on-off control are similar to those of phase control and the performance analysis is also similar. For these reasons, the phase-control techniques are only discussed and analyzed in this chapter.

Example 11.1 Finding the Performances of an Ac Voltage Controller with On-Off Control

An ac voltage controller in Figure 11.1a has a resistive load of $R = 10 \, \Omega$ and the root-mean-square (rms) input voltage is $V_s = 120 \, V$, 60 Hz. The thyristors switch is on for $n = 25$ cycles and is off for $m = 75$ cycles. Determine (a) the rms output voltage $V_o$, (b) the input power factor (PF), and (c) the average and rms current of thyristors.

Solution

$R = 10 \, \Omega$, $V_s = 120 \, V$, $V_m = \sqrt{2} \times 120 = 169.7 \, V$, and $k = n/(m + n) = 25/100 = 0.25$.

a. From Eq. (11.1), the rms value of output voltage is

$$V_o = V_s \sqrt{k} = V_s \sqrt{\frac{n}{m + n}} = 120 \sqrt{\frac{25}{100}} = 60 \, V$$

and the rms load current is $I_o = V_o/R = 60/10 = 6.0 \, A$.

b. The load power is $P_o = I_o^2 R = 6^2 \times 10 = 360 \, W$. Because the input current is the same as the load current, the volt-ampere (VA) input is

$$VA = V_s I_s = V_s I_o = 120 \times 6 = 720 \, W$$

The input PF is

$$PF = \frac{P_o}{VA} = \sqrt{\frac{n}{m + n}} = \sqrt{k}$$

$$= \sqrt{0.25} = \sqrt{\frac{360}{720}} = 0.5 \text{ (lagging)}$$  \hspace{1cm} (11.2)

P.J. Shah
11.3 Principle of Phase Control

The peak thyristor current is \( I_m = V_m/R = 169.7/10 = 16.97 \text{ A} \). The average current of thyristors is

\[
I_A = \frac{n}{2\pi(m + n)} \int_0^\pi I_m \sin \omega t \, d(\omega t) = \frac{I_m n}{\pi(m + n)} = \frac{kl_m}{n}
\]

\[
= \frac{16.97}{\pi} \times 0.25 = 1.33 \text{ A}
\]

The rms current of thyristors is

\[
I_R = \left( \frac{n}{2\pi(m + n)} \int_0^n I_m \sin^2 \omega t \, d(\omega t) \right)^{1/2} = \frac{I_m}{2} \sqrt{\frac{n}{m + n}} = \frac{I_m \sqrt{k}}{2}
\]

\[
= \frac{16.97}{2} \times \sqrt{0.25} = 4.24 \text{ A}
\]

Notes:

1. The PF and output voltage vary with the square root of the duty cycle. The PF is poor at the low value of the duty cycle \( k \) and is shown in Figure 11.1c.

2. If \( T \) is the period of the input voltage, \( (m + n)T \) is the period of on–off control. \( (m + n)T \) should be less than the mechanical or thermal time constant of the load, and is usually less than 1 s, but not in hours or days. The sum of \( m \) and \( n \) is generally around 100.

3. If Eq. (11.2) is used to determine the PF with \( m \) and \( n \) in days, it can give erroneous results. For example, if \( m = 3 \) days and \( n = 3 \) days, Eq. (11.2) gives PF = \( 3/(3 + 3) \)^1/2 = 0.707, which is not physically possible. Because if the controller is on for 3 days and off for 3 days, the PF becomes independent on the load impedance angle \( \theta \).

Key Points of Section 11.2

- The on-off control switches the ac supply to the load for an integral number \( m \) of supply-frequency cycles and then disconnect the load for a certain number \( n \) of supply cycles. This type of control is applied in applications that have a high mechanical inertia and high thermal constant.

11.3 PRINCIPLE OF PHASE CONTROL

The principle of phase control can be explained with reference to Figure 11.2a. The power flow to the load is controlled by delaying the firing angle of thyristor \( T_1 \). Figure 11.2b illustrates the gate pulses of thyristor \( T_1 \) and the waveforms for the input and output voltages. Due to the presence of diode \( D_1 \), the control range is limited and the effective rms output voltage can only be varied between 70.7 and 100%. The output voltage and input current are asymmetric and contain a dc component. If there is an
input transformer, it may cause a saturation problem. This circuit is a single-phase half-wave controller and is suitable only for low-power resistive loads, such as heating and lighting. Because the power flow is controlled during the positive half-cycle of input voltage, this type of controller is also known as a unidirectional controller.

If \( v_s = V_m \sin \omega t = \sqrt{2} V_s \sin \omega t \) is the input voltage and the delay angle of thyristor \( T_1 \) is \( \omega t = \alpha \), the rms output voltage is found from

\[
V_o = \left\{ \frac{1}{2\pi} \left[ \int_0^\pi 2V_s^2 \sin^2 \omega t \, d(\omega t) + \int_\pi^{2\pi} 2V_s^2 \sin^2 \omega t \, d(\omega t) \right] \right\}^{1/2}
\]

\[
= \left\{ \frac{2V_s^2}{4\pi} \left[ \int_0^\pi (1 - \cos 2\omega t) \, d(\omega t) + \int_\pi^{2\pi} (1 - \cos 2\omega t) \, d(\omega t) \right] \right\}^{1/2} \quad (11.5)
\]

\[
= V_s \left[ \frac{1}{2\pi} \left( 2\pi - \alpha + \frac{\sin 2\alpha}{2} \right) \right]^{1/2}
\]

The average value of output voltage is

\[
V_{dc} = \frac{1}{2\pi} \left[ \int_0^\pi \sqrt{2} V_s \sin \omega t \, d(\omega t) + \int_\pi^{2\pi} \sqrt{2} V_s \sin \omega t \, d(\omega t) \right]
\]

\[
= \frac{\sqrt{2}}{2\pi} V_s \cos \alpha - 1 \quad (11.6)
\]

If \( \alpha \) is varied from 0 to \( \pi \), \( V_o \) varies from \( V_s \) to \( V_s/\sqrt{2} \) and \( V_{dc} \) varies from 0 to \( -\sqrt{2} V_s/\pi \).

**Gating sequence.** The gating sequence is as follows:

1. Generate a pulse signal at the positive zero crossing of the supply voltage \( v_s \).
2. Delay the pulse by the desired angle \( \alpha \) and apply it between the gate and source terminals of \( T_1 \) through a gate-isolating circuit.
Example 11.2 Finding the Performance Parameters of a Single-Phase Half-Wave Controller

A single-phase ac voltage controller in Figure 11.2a has a resistive load of \( R = 10 \, \Omega \) and the input voltage is \( V_s = 120 \, V \), 60 Hz. The delay angle of thyristor \( T_1 \) is \( \alpha = \pi/2 \). Determine (a) the rms value of output voltage \( V_o \), (b) the input PF, and (c) the average input current.

**Solution**

\( R = 10 \, \Omega \), \( V_s = 120 \, V \), \( \alpha = \pi/2 \), and \( V_m = \sqrt{2} \times 120 = 169.7 \, V \).

a. From Eq. (6-5), the rms value of the output voltage

\[
V_o = 120 \sqrt{3} \frac{1}{4} = 103.92 \, V
\]

b. The rms load current

\[
I_o = \frac{V_o}{R} = \frac{103.92}{10} = 10.392 \, A
\]

The load power

\[
P_o = I_o^2 R = 10.392^2 \times 10 = 1079.94 \, W
\]

Because the input current is the same as the load current, the input VA rating is

\[
VA = V_s I_s = V_V I_v = 120 \times 10.392 = 1247.04 \, VA
\]

The input PF

\[
PF = \frac{P_o}{VA} = \frac{V_o}{V_s} = \sqrt{3} \frac{1}{2 \pi} \left( \frac{2 \pi - \alpha + \frac{\sin 2\alpha}{2}}{2} \right)^{1/2} = \frac{1079.94}{1247.04} = 0.866 \text{ (lagging)}
\]

(11.7)

c. From Eq. (11.6), the average output voltage

\[
V_{dc} = -120 \times \frac{\sqrt{2}}{2\pi} = -27 \, V
\]

and the average input current

\[
I_D = \frac{V_{dc}}{R} = \frac{-27}{10} = -2.7 \, A
\]

**Note:** The negative sign of \( I_D \) signifies that the input current during the positive half-cycle is less than that during the negative half-cycle. If there is an input transformer, the transformer core may be saturated. The unidirectional control is not normally used in practice. However, it explains the principle of the phase control for ac voltage controllers.

**Key Points of Section 11.3**

- Although the half-wave controller can vary the output voltage by varying the delay angle \( \alpha \), the output contains an undesirable dc component.
- This type of controller is not generally used in practical applications.
11.4 SINGLE-PHASE BIDIRECTIONAL CONTROLLERS WITH RESISTIVE LOADS

The problem of dc input current can be prevented by using bidirectional (or full-wave) control, and a single-phase full-wave controller with a resistive load is shown in Figure 11.3a. During the positive half-cycle of input voltage, the power flow is controlled by varying the delay angle of thyristor $T_1$; and thyristor $T_2$ controls the power flow during the negative half-cycle of input voltage. The firing pulses of $T_1$ and $T_2$ are kept 180° apart. The waveforms for the input voltage, output voltage, and gating signals for $T_1$ and $T_2$ are shown in Figure 11.3b.

If $v_s = \sqrt{2} V_s \sin \omega t$ is the input voltage, and the delay angles of thyristors $T_1$ and $T_2$ are equal ($\alpha_2 = \pi + \alpha_1$), the rms output voltage can be found from

$$V_o = \left\{ \frac{2}{2\pi} \int_0^\pi 2V_s^2 \sin^2 \omega t \, d(\omega t) \right\}^{1/2}$$

$$= \left\{ \frac{4V_s^2}{4\pi} \int_0^\pi (1 - \cos 2\omega t) \, d(\omega t) \right\}^{1/2}$$

$$= V_s \left[ \frac{1}{\pi} \left( \pi - \alpha + \frac{\sin 2\alpha}{2} \right) \right]^{1/2}$$

(11.8)

By varying $\alpha$ from 0 to $\pi$, $V_o$ can be varied from $V_s$ to 0.

In Figure 11.3a, the gating circuits for thyristors $T_1$ and $T_2$ must be isolated. It is possible to have a common cathode for $T_1$ and $T_2$ by adding two diodes, as shown in Figure 11.4. Thyristor $T_1$ and diode $D_1$ conduct together during the positive half-cycle;
11.4 Single-Phase Bidirectional Controllers with Resistive Loads

![Circuit Diagram](image)

**FIGURE 11.4**
Single-phase full-wave controller with common cathode.

and thyristor T₂ and diode D₂ conduct during the negative half-cycle. Because this circuit can have a common terminal for gating signals of T₁ and T₂, only one isolation circuit is required, but at the expense of two power diodes. Due to two power devices conducting at the same time, the conduction losses of devices would increase and efficiency would be reduced.

A single-phase full-wave controller can also be implemented with one thyristor and four diodes, as shown in Figure 11.5a. The four diodes act as a bridge rectifier. The voltage across thyristor T₁ and its current are always unidirectional. With a resistive load, the thyristor current would fall to zero due to natural commutation in every half-cycle, as shown in Figure 11.5b. However, if there is a large inductance in the circuit, thyristor T₁ may not be turned off in every half-cycle of input voltage, and this may result in a loss of control. It would require detecting the zero crossing of the load current to guarantee turn-off of the conducting thyristor before firing the next one. Three power devices conduct at the same time and the efficiency is also reduced. The bridge

![Circuit Diagram](image)

**FIGURE 11.5**
Single-phase full-wave controller with one thyristor.
rectifier and thyristor (or transistor) act as a \textit{bidirectional switch}, which is commercially available as a single device with a relatively low on-state conduction loss.

\textbf{Gating sequence.} The gating sequence is as follows:

1. Generate a pulse signal at the positive zero crossing of the supply voltage $v_s$.
2. Delay the pulse by the desired angle $\alpha$ for gating $T_1$ through a gate-isolating circuit.
3. Generate another pulse of delay angle $\alpha + \pi$ for gating $T_2$.

\textbf{Example 11.3 Finding the Performance Parameters of a Single-Phase Full-Wave Controller}

A single-phase full-wave ac voltage controller in Figure 11.3a has a resistive load of $R = 10 \ \Omega$ and the input voltage is $V_r = 120 \ \text{V (rms)}$, 60 Hz. The delay angles of thyristors $T_1$ and $T_2$ are equal: $\alpha_1 = \alpha_2 = \alpha = \pi/2$. Determine (a) the rms output voltage $V_o$, (b) the input PF, (c) the average current of thyristors $I_A$, and (d) the rms current of thyristors $I_R$.

\textit{Solution}

$a$. From Eq. (11.8), the rms output voltage

$$V_o = \frac{120}{\sqrt{2}} = 84.85 \ \text{V}$$

$b$. The rms value of load current is $I_o = V_o/R = 84.85/10 = 8.485 \ \text{A}$ and the load power is $P_o = I_o^2R = 8.485^2 \times 10 = 719.95 \ \text{W}$. Because the input current is the same as the load current, the input VA rating is

$$VA = V_i I_i = V_o I_o = 120 \times 8.485 = 1018.2 \ \text{W}$$

The input PF is

$$PF = \frac{P_o}{VA} = \frac{V_o}{V_r} = \left[\frac{1}{\pi} \left( \pi - \alpha + \frac{\sin 2\alpha}{2} \right) \right]^{1/2}$$

$$= \frac{1}{\sqrt{2}} = \frac{719.95}{1018.2} = 0.707 \ \text{(lagging)}$$

$c$. The average thyristor current

$$I_A = \frac{1}{2\pi R} \int_{-\alpha}^{\alpha} \sqrt{2} V_i \sin \omega t \ d(\omega t)$$

$$= \frac{\sqrt{2} V_o}{2\pi R} (\cos \alpha + 1)$$

$$= \sqrt{2} \times \frac{120}{2\pi \times 10} = 2.7 \ \text{A}$$
d. The rms value of the thyristor current

\[
I_R = \left[ \frac{1}{2\pi R^2} \int_{\alpha}^{\pi} 2V_s^2 \sin^2 \omega t \, d(\omega t) \right]^{1/2} \\
= \left[ \frac{2V_s^2}{4\pi R^2} \int_{\alpha}^{\pi} (1 - \cos 2\omega t) \, d(\omega t) \right]^{1/2} \\
= \frac{V_s}{\sqrt{2} R} \left[ \frac{1}{\pi} \left( \pi - \alpha + \frac{\sin 2\alpha}{2} \right) \right]^{1/2} \\
= \frac{120}{2 \times 10} = 6A
\]

Key Points of Section 11.4

- Varying the delay angle \( \alpha \) from 0 to \( \pi \) can vary the rms output voltage from \( V_s \) to 0.
- The output of this controller contains no dc component.

11.5 SINGLE-PHASE CONTROLLERS WITH INDUCTIVE LOADS

Section 11.4 deals with the single-phase controllers with resistive loads. In practice, most loads are inductive to a certain extent. A full-wave controller with an \( RL \) load is shown in Figure 11.6a. Let us assume that thyristor \( T_1 \) is fired during the positive half-cycle and carries the load current. Due to inductance in the circuit, the current of thyristor \( T_1 \) would not fall to zero at \( \omega t = \pi \), when the input voltage starts to be negative. Thyristor \( T_1 \) continues to conduct until its current \( i_1 \) falls to zero at \( \omega t = \beta \). The conduction angle of thyristor \( T_1 \) is \( \delta = \beta - \alpha \) and depends on the delay angle \( \alpha \) and the PF angle of load \( \theta \). The waveforms for the thyristor current, gating pulses, and input voltage are shown in Figure 11.6b.

If \( v_i = \sqrt{2} V_s \sin \omega t \) is the instantaneous input voltage and the delay angle of thyristor \( T_1 \) is \( \alpha \), thyristor current \( i_1 \) can be found from

\[
L \frac{di_1}{dt} + R i_1 = \sqrt{2} V_s \sin \omega t
\]  
(11.12)

The solution of Eq. (11.12) is of the form

\[
i_1 = \frac{\sqrt{2} V_s}{Z} \sin(\omega t - \theta) + A_1 e^{-(R/L)t}
\]  
(11.13)

where load impedance \( Z = [R^2 + (\omega L)^2]^{1/2} \) and load angle \( \theta = \tan^{-1} (\omega L/R) \).

The constant \( A_1 \) can be determined from the initial condition: at \( \omega t = \alpha \), \( i_1 = 0 \).

From Eq. (11.13) \( A_1 \) is found as

\[
A_1 = -\frac{\sqrt{2} V_s}{Z} \sin(\alpha - \theta) e^{(R/L)(\omega \alpha)}
\]  
(11.14)

Substitution of \( A_1 \) from Eq. (11.14) in Eq. (11.13) yields

\[
i_1 = \frac{\sqrt{2} V_s}{Z} [\sin(\omega t - \theta) - \sin(\alpha - \theta) e^{(R/L)(\omega \alpha - \omega \alpha)}]
\]  
(11.15)
The angle $\beta$, when current $i_1$ falls to zero and thyristor $T_1$ is turned off, can be found from the condition $i_1(\omega t = \beta) = 0$ in Eq. (11.15) and is given by the relation

$$\sin(\beta - \theta) = \sin(\alpha - \theta)e^{(R/L)(\alpha-\beta)/\omega}$$  \hspace{1cm} (11.16)

The angle $\beta$, which is also known as an extinction angle, can be determined from this transcendental equation and it requires an iterative method of solution. Once $\beta$ is known, the conduction angle $\delta$ of thyristor $T_1$ can be found from

$$\delta = \beta - \alpha$$  \hspace{1cm} (11.17)

The rms output voltage

$$V_o = \left[ \frac{2}{2\pi} \int_\alpha^\beta 2V_i^2 \sin^2 \omega t \, dt(\omega t) \right]^{1/2}$$
\[ I_R = \left[ \frac{1}{\pi} \int_{\alpha}^{\beta} i_1^2 \, d(\omega t) \right]^{1/2} \]

\[ = \frac{V_v}{Z} \left[ \frac{1}{\pi} \int_{\alpha}^{\beta} \left\{ \sin(\omega t - \theta) - \sin(\alpha - \theta) e^{\frac{iRL}{L}(\omega t - \theta)} \right\}^2 \, d(\omega t) \right]^{1/2} \]  

(11.19)

The rms output current can then be determined by combining the rms current of each thyristor as

\[ I_0 = \sqrt{I_R^2 + I_R^2} = \sqrt{2} I_R \]  

(11.20)

The average value of thyristor current can also be found from Eq. (11.15) as

\[ I_A = \frac{1}{2\pi} \int_{\alpha}^{\beta} i_1 \, d(\omega t) \]

\[ = \frac{\sqrt{2} V_v}{2\pi Z} \int_{\alpha}^{\beta} \left[ \sin(\omega t - \theta) - \sin(\alpha - \theta) e^{\frac{iRL}{L}(\omega t - \theta)} \right] \, d(\omega t) \]  

(11.21)

The gating signals of thyristors could be short pulses for a controller with resistive loads. However, such short pulses are not suitable for inductive loads. This can be explained with reference to Figure 11.6b. When thyristor \( T_2 \) is fired at \( \omega t = \pi + \alpha \), thyristor \( T_1 \) is still conducting due to load inductance. By the time the current of thyristor \( T_1 \) falls to zero and \( T_1 \) is turned off at \( \omega t = \beta = \alpha + 8 \), the gate pulse of thyristor \( T_2 \) has already ceased and consequently, \( T_2 \) cannot be turned on. As a result, only thyristor \( T_1 \) operates, causing asymmetric waveforms of output voltage and current. This difficulty can be resolved by using continuous gate signals with a duration of \( (\pi - \alpha) \), as shown in Figure 11.6c. As soon as the current of \( T_1 \) falls to zero, thyristor \( T_2 \) (with gate pulses as shown in Figure 11.6c) would be turned on. However, a continuous gate pulse increases the switching loss of thyristors and requires a larger isolating transformer for the gating circuit. In practice, a train of pulses with short durations as shown in Figure 11.6d are normally used to overcome these problems.

The waveforms for the output voltage \( V_0 \), output current \( I_0 \), and the voltage across \( T_1, v_T1 \) are shown in Figure 11.7 for an \( RL \) load. There may be a short hold-off angle \( \gamma \) after the zero crossing of negative going current.

Equation (11.15) indicates that the load voltage (and current) can be sinusoidal if the delay angle \( \alpha \) is less than the load angle \( \theta \). If \( \alpha \) is greater than \( \theta \), the load current would be discontinuous and nonsinusoidal.
FIGURE 11.7
Typical waveforms of single-phase ac voltage controller with an RL load.

Notes:

1. If $\alpha = \theta$, from Eq. (11.16),

$$\sin(\beta - \theta) = \sin(\beta - \alpha) = 0$$

(11.22)

and

$$\beta - \alpha = \delta = \pi$$

(11.23)

2. Because the conduction angle $\delta$ cannot exceed $\pi$ and the load current must pass through zero, the delay angle $\alpha$ may not be less than $\theta$ and the control range of delay angle is

$$\theta \leq \alpha \leq \pi$$

(11.24)

3. If $\alpha \leq \theta$ and the gate pulses of thyristors are of long duration, the load current would not change with $\alpha$, but both thyristors would conduct for $\pi$. Thyristor $T_1$ would turn on at $\omega t = \theta$ and thyristor $T_2$ would turn off at $\omega t = \pi + \theta$. 

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11.5 Single-Phase Controllers with Inductive Loads

Gating sequence. The gating sequence is as follows:

1. Generate a train of pulse signal at the positive zero crossing of the supply voltage $v_i$ [1].
2. Delay this pulse by the desired angle $\alpha$ for gating $T_1$ through a gate-isolating circuit.
3. Generate another continuous pulse of delay angle $\alpha + \pi$ for gating.

Example 11.4 Finding the Performance Parameters of a Single-Phase Full-Wave Controller with an RL Load

The single-phase full-wave controller in Figure 11.6a supplies an RL load. The input rms voltage is $V_i = 120$ V, 60 Hz. The load is such that $L = 6.5$ mH and $R = 2.5$ $\Omega$. The delay angles of thyristors are equal: $\alpha_1 = \alpha_2 = \pi/2$. Determine (a) the conduction angle of thyristor $T_1$, $\theta$; (b) the rms output voltage $V_o$; (c) the rms thyristor current $I_R$; (d) the rms output current $I_o$; (e) the average current of a thyristor $I_{av}$; and (f) the input PF.

Solution

$R = 2.5$ $\Omega$, $L = 6.5$ mH, $f = 60$ Hz, $\omega = 2\pi \times 60 = 377$ rad/s, $V_i = 120$ V, $\alpha = 90^\circ$, and $\theta = \tan^{-1}(\omega L/R) = 44.43^\circ$.

a. The extinction angle can be determined from the solution of Eq. (11.16) and an iterative solution yields $\beta = 220.35^\circ$. The conduction angle is $\delta = \beta - \alpha = 220.43 - 90 = 130.43^\circ$.

b. From Eq. (11.18), the rms output voltage is $V_o = 68.09$ V.

c. Numerical integration of Eq. (11.19) between the limits $\omega t = \alpha$ to $\beta$ gives the rms thyristor current as $I_R = 15.07$ A.

d. From Eq. (11.20), $I_o = \sqrt{2} \times 15.07 = 21.3$ A.

e. Numerical integration of Eq. (11.21) yields the average thyristor current as $I_{av} = 8.23$ A.

f. The output power $P_o = 21.3^2 \times 2.5 = 1134.2$ W, and the input VA rating is $VA = 120 \times 21.3 = 2556$ W; therefore,

\[
PF = \frac{P_o}{VA} = \frac{1134.200}{2556} = 0.444 \text{ (lagging)}
\]

Note: The switching action of thyristors makes the equations for currents nonlinear. A numerical method of solution for the thyristor conduction angle and currents is more efficient than classical techniques. A computer program is used to solve this example. Students are encouraged to verify the results of this example and to appreciate the usefulness of a numerical solution, especially in solving nonlinear equations of thyristor circuits.

Key Points of Section 11.5

- An inductive load extends the load current beyond $\pi$. The load current can be continuous if the delay angle $\alpha$ is less than the impedance angle $\theta$.
- For $\alpha > 0$, which is usually the case, the load current is discontinuous. Thus, the control range is $0 \leq \alpha \leq \pi$. 

11.6 THREE-PHASE FULL-WAVE CONTROLLERS

The unidirectional controllers, which contain dc input current and higher harmonic content due to the asymmetric nature of the output voltage waveform, are not normally used in ac motor drives; a three-phase bidirectional control is commonly used. The circuit diagram of a three-phase full-wave (or bidirectional) controller is shown in Figure 11.8 with a Y-connected resistive load. The firing sequence of thyristors is $T_1, T_2, T_3, T_4, T_5, T_6$.

If we define the instantaneous input phase voltages as

$$v_{AN} = \sqrt{2} V_s \sin \omega t$$
$$v_{BN} = \sqrt{2} V_s \sin \left( \omega t - \frac{2\pi}{3} \right)$$
$$v_{CN} = \sqrt{2} V_s \sin \left( \omega t - \frac{4\pi}{3} \right)$$

the instantaneous input line voltages are

$$v_{AB} = \sqrt{6} V_s \sin \left( \omega t + \frac{\pi}{6} \right)$$
$$v_{BC} = \sqrt{6} V_s \sin \left( \omega t - \frac{\pi}{2} \right)$$
$$v_{CA} = \sqrt{6} V_s \sin \left( \omega t - \frac{7\pi}{6} \right)$$

The waveforms for the input voltages, conduction angles of thyristors, and output phase voltages are shown in Figure 11.9 for $\alpha = 60^\circ$ and $\alpha = 120^\circ$. For $0 \leq \alpha < 60^\circ$, 

![Diagram of three-phase bidirectional controller](image-url)
immediately before the firing of \( T_1 \), two thyristors conduct. Once \( T_1 \) is fired, three thyristors conduct. A thyristor turns off when its current attempts to reverse. The conditions alternate between two and three conducting thyristors.

For \( 60^\circ \leq \alpha < 90^\circ \), only two thyristors conduct at any time. For \( 90^\circ \leq \alpha < 150^\circ \), although two thyristors conduct at any time, there are periods when no thyristors are on. For \( \alpha \geq 150^\circ \), there is no period for two conducting thyristors and the output voltage becomes zero at \( \alpha = 150^\circ \). The range of delay angle is

\[
0 \leq \alpha \leq 150^\circ
\]

Similar to half-wave controllers, the expression for the rms output phase voltage depends on the range of delay angles. The rms output voltage for a Y-connected load
6

Application

Objectives
To study various power electronics applications like:

- Uninterruptible power supply (UPS), its types, battery selection and its charging.
- Electronic ballast and its advantages over conventional ballast.
- Separately excited DC motor drive.
- High voltage DC and AC transmission, various converters and arrangements required for it.
- High frequency induction heating and electric welding.

6.1 Uninterruptible Power Supply (UPS)

- An UPS is used to provide the power when mains is not available. In the present days of load shading, UPS is playing major role. UPS is being used along with computers.

- There are three types of UPS as follows
  i) Online UPS
  ii) Offline UPS
  iii) Line interactive UPS

The block diagrams and working of these UPS is discussed next.

6.1.1 Online UPS

The online UPS is also called inverter preferred UPS. Fig. 6.1.1 shows the block diagram of online UPS.
6.1.1 Block diagram of online UPS system

- When the main supply is present, the rectifier/charger provides power to an inverter as well as battery (See Fig. 6.1.2(a)). The battery is charged. The inverter is on and feeds power to the load through UPS static switch.
- The UPS static switch is always on and connects load to inverter output.
- The mains static switch is always off. But when the UPS fails, then load is connected directly to the mains directly through mains static switch.
- When the mains supply is not available, then battery bank supplies power to an inverter (See Fig. 6.1.2(b)). Thus an inverter is always on and it takes power from rectifier or battery...
- Fig. 6.1.2 shows the power flow when mains is present and mains is absent.

6.1.2 Offline UPS

The offline UPS is also called line preferred UPS. Fig. 6.1.3 shows the block diagram of an offline UPS. Observe that this diagram appears similar to that of online UPS, but it is functionally different.
Fig. 6.1.3 Block diagram of offline UPS

- When mains supply is present, then charger charges the battery. Inverter is off and UPS static switch is off. The load is connected to mains through mains static switch. The power flow is shown in Fig. 6.1.4(a).
- When mains supply is not available, then inverter is turned on.
- Inverter takes power from the battery. The load is connected to inverter output through UPS static switch. The power flow diagram is shown in Fig. 6.1.4(b).

Fig. 6.1.4 Power flow in offline UPS

- The mains static switch is always on and keeps load connected to mains. The mains static switch is turned off when mains is not available.
- The charger feeds power only to the battery. Hence its power handling capacity is reduced.
6.1.3 Line Interactive UPS

Fig. 6.1.5 shows the block diagram of line interactive UPS system:

- When the mains supply is present, the static switch is 'on'. The static switch connects load to mains supply through inductor L. The batteries are charged through the charger block. Fig. 6.1.6(a) shows the power flow diagram.

6.1.6 Power flow in line interactive UPS

- When mains is absent, the mains static switch is open. The inverter then turns on and supplies power to the load.
- The charger/inverter block operates as charger when mains is available and as an inverter when mains is not available.

6.1.4 Selection of the Battery and AH Rating

While designing UPS system, selection of the battery is critical part. The battery has to provide estimated backup and it should be charged properly.
Power Electronics 6 - 5 Applications

AH rating: AH rating of the battery tells about how much amount of current it can supply for one hour. For example 80 AH means the battery will supply 80A for 1 hour.

Efficiency of the battery

i) AH efficiency: An ampere-hour efficiency is defined as the ratio of AH taken from the battery while discharging to AH given to the battery while charging i.e.

\[
\text{AH efficiency} = \frac{\text{AH (discharging)}}{\text{AH (charging)}} \quad \text{...(6.1.1)}
\]

AH efficiency varies from 90 to 95%.

ii) WH efficiency: WH efficiency is the ratio of energy obtained from battery while discharging to energy given to the battery while charging i.e.,

\[
\text{WH efficiency} = \frac{\text{Energy (discharging)}}{\text{Energy (charging)}} \quad \text{...(6.1.2)}
\]

\[
= \frac{\text{AH (discharging)} \times \text{Average voltage (discharging)}}{\text{AH (charging)} \times \text{Average voltage (charging)}} \quad \text{...(6.1.3)}
\]

\[
= \text{AH efficiency} \times \frac{\text{Average voltage on discharging}}{\text{Average voltage on charging}} \quad \text{...(6.1.4)}
\]

WH efficiency varies from 70 to 80%.

Selection of the battery

The capacity of the battery is given as,

\[
\text{Battery kW} = \frac{\text{Load kVA} \times \text{Power factor}}{\text{Inverter efficiency}} \quad \text{...(6.1.5)}
\]

Example 6.1.1: An online UPS is driving 800 W, 0.8 lagging PF load, an inverter efficiency is 80% and dc link voltage and battery voltage is 48 V dc. Assuming batteries are ideal,

Find i) VA rating of an inverter

ii) Wattage or peak power requirement of rectifier.

iii) AH capacity of batteries required for backup time of 30 minutes.

[May-2000, 8 Marks]

Solution: Given data

\[
P_{0(\text{UPS})} = 800 \text{ W}, \quad \text{PF} = 0.8
\]

\[
\eta_{(\text{inverter})} = 0.8, \quad V_{dc} = 48 \text{ V}
\]
i) To obtain VA rating of an inverter

\[ PF = \frac{\text{Active power output}}{\text{Total rms power (kVA)}} \]

\[ \text{Total rms power (kVA)} = \frac{\text{Active power output}}{PF} \]

\[ = \frac{800}{0.8} \quad \text{Here active power is } P_{\text{in(Hrs)}} \]

\[ = 1000 \text{ VA} \]

ii) To obtain wattage of rectifier

Assuming that there is separate battery charger and rectifier supplies power only to an inverter, then

\[ \text{Rectifier wattage} = \frac{\text{Active power supplied by inverter}}{\text{Inverter efficiency}} \]

\[ = \frac{P_{\text{in(Hrs)}}}{\eta_{\text{(inverter)}}} = \frac{800}{0.8} = 1000 \text{ W} \]

iii) To obtain AH capacity

Here rectifier wattage is 1000 W. When mains fails, the battery must supply 1000 W to an inverter.

\[ \text{Battery wattage} = \text{Battery voltage} \times \text{Battery current} \]

\[ = V_{dc} \times I_{dc} \]

\[ 1000 = 48 \times I_{dc} \]

\[ I_{dc} = 20.83 \]

Since the battery is ideal, the AH rating for 30 minutes (\( \frac{1}{2} \text{Hr} \)) backup will be,

\[ \text{AH rating} = I_{dc} \times \text{Backup(Hrs)} \]

\[ = 20.83 \times \frac{1}{2} = 10.41 \text{ AH} \]

\[ \approx 11 \text{ AH} \]

Example 6.1.2: A 1 kVA, 230 V, 50 Hz UPS feeds a 0.8 power factor load and operates from 96 V DC bus, the inverter efficiency being 90%. Calculate backup time available if battery has the capacity of 100 AH. Also calculate the peak output power of
input rectifier cum battery charger if batteries must be restored with six hours of mains supply becoming available. Assume the following:

i) battery float voltage is 110 V

ii) battery voltage at start of discharge is 100 V.

iii) battery voltage at end of discharge is 88 V.

iv) capacity derating factor of battery upto 10 hour discharge is 0.5.

v) battery voltage during charging is 106 V.

vi) battery charging efficiency is 75%.

[May-2004, May-2005, 8 Marks]

Ans. : Given : UPS ratings : 1 kVA 230 V, 50 Hz

Load PF = 0.8, \( V_{dc} = 96 \text{ V} \)

\[ \eta_{\text{inverter}} = 0.9, \text{AH rating} = 100 \]

Charging time = 6 Hrs.

\[ \text{Battery kW} = \frac{\text{Load kVA} \times \text{PF}}{\text{Inverter efficiency}} \]

\[ = \frac{1 \text{ kVA} \times 0.8}{0.9} = 0.8889 \text{ kW} \]

Battery voltage at the end of discharge is 88 V. Hence the discharge current will be maximum at the end of discharge i.e.

\[ \text{Battery kW} = \text{Battery voltage} \times I_{dc} \]

\[ 0.8889 \times 10^3 = 88 \times I_{dc} \]

\[ I_{dc} = 10.10 \text{ A} \]

The capacity derating factor is given as 0.5. This means even though the capacity of the battery is 100 AH, it should be discharged with 50 AH capacity. Therefore backup time will be,

\[ \text{Backup time} = \frac{\text{AH capacity (after derating)}}{I_{dc}} \]

\[ = \frac{50}{10.10} = 5 \text{ Hrs.} \]

To obtain rectifier cum charger output power

Battery charging efficiency = \[ \frac{\text{AH output}}{\text{AH input}} \]
\[
0.75 = \frac{100 \text{AH}}{\text{AH input}}
\]

\[\therefore \quad \text{All input} = 133.33 \text{ AH.}\]

\[\text{Charging current} = \frac{\text{AH input}}{\text{Charging time}} = \frac{133.33}{6} = 22.22 \text{ A.}\]

The rectifier cum charger has to supply current to the inverter as well as battery in case of online UPS. The input kW to the inverter is 0.8889 kW. The DC voltage during charging is 106 V.

Therefore the input current to inverter is,
\[I_{dc} = \frac{0.8889 \text{ kW}}{106 \text{ V}} = 8.3858 \text{ A}.\]

Therefore the total current that the rectifier cum charger that must supply is,

\[
\text{Rectifier current} = \text{Battery charging current} + \text{Inverter input current}
\]
\[= 22.22 \text{ A} + 8.3858 \text{ A}
\]
\[= 30.6 \text{ A}.
\]

\[\therefore \quad \text{Peak output power of rectifier} = \text{Rectifier voltage} \times \text{Rectifier current}
\]
\[= 106 \times 30.6
\]
\[= 3244.22 \text{ Watts}
\]

### 6.1.5 Comparison between Online and Offline UPS


**Advantages of online UPS**

i) It provides isolation between main supply and load.

ii) Since inverter is always on, the quality of load voltage is free from distortion.

iii) All the disturbances of supply such as blackout, brownout, spikes etc. are absent in the output.

iv) Voltage regulation is better.

v) Transfer time is practically zero since inverter is always on.

**Disadvantages of online UPS**

i) Overall efficiency of UPS is reduced since inverter is always on.

ii) The wattage of the rectifier is increased since it has to supply power to inverter as well as charge battery.

iii) Online UPS is costlier than other UPS systems.
Applications of online UPS
  i) Induction motor drives and similar other motor control applications.
  ii) Intensive care units, medical equipments.

Advantages of offline UPS
  i) Offline UPS has high efficiencies, since charger is not continuously on.
  ii) The power handling capacity of charger is reduced.
  iii) Offline UPS are not very costly.
  iv) Internal control is simpler in offline UPS.

Disadvantages of offline UPS
  i) Since offline UPS provides mains supply when it is present, the output contains voltage spikes, brownouts, blackouts.
  ii) There is finite transfer time from mains to inverter when mains supply fails.
  iii) Output of offline UPS is not perfectly reliable.

Applications of offline UPS
  i) Computers, printers, scanners etc use offline UPS.
  ii) Emergency power supplies, EPABX.

<table>
<thead>
<tr>
<th>Sr. No.</th>
<th>Parameter</th>
<th>Online UPS</th>
<th>Offline UPS</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Inverter</td>
<td>Always on</td>
<td>Turned on when mains fails.</td>
</tr>
<tr>
<td>2.</td>
<td>Rectifier cum charger</td>
<td>Supplies power to inverter as well as charges battery</td>
<td>Charges only battery.</td>
</tr>
<tr>
<td>3.</td>
<td>Output waveform</td>
<td>sine wave</td>
<td>Quasi square wave</td>
</tr>
<tr>
<td>4.</td>
<td>Harmonic distortion</td>
<td>Low</td>
<td>High</td>
</tr>
<tr>
<td>5.</td>
<td>Efficiency</td>
<td>Low</td>
<td>High</td>
</tr>
<tr>
<td>6.</td>
<td>Load</td>
<td>Isolated from supply</td>
<td>Not isolated from supply</td>
</tr>
<tr>
<td>7.</td>
<td>Cost</td>
<td>High</td>
<td>Low</td>
</tr>
</tbody>
</table>

Table 6.1.1 Comparison between online and offline UPS

Review Questions
1. Explain the operation of offline UPS with the help of block diagram.
2. Compare offline and online UPS.
3. How battery is selected for the UPS system?
4. State the advantages and disadvantages of online UPS over offline UPS.
6.4.2 1φ Semiconverter Based DC Motor Drive

Fig. 6.4.2 shows the circuit diagram having 1φ semiconverter for the armature voltage control.

In the above circuit observe that, the field supply is given by uncontrolled rectifier. Hence field flux is fixed. The armature voltage can be varied by controlling the firing angles of $T_1$ and $T_2$ in semiconverter. The load current can be continuous or discontinuous depending upon armature inductance.

6.4.2.1 Continuous Load Current Mode

The armature current is continuous in this mode. Fig. 6.4.3 shows the waveforms for this mode.

The supply voltage and the back emf is shown in Fig. 6.4.3 (a). In Fig. 6.4.3 (b) observe that the output voltage is zero from $\pi$ to $\pi+\alpha$ due to free wheeling operation. Fig. 6.4.3 (c) shows that the armature current is continuous. The ripple in this current depends upon the armature inductance of the motor.

Fig. 6.4.3 (d) shows the supply current $i_s$ observe that supply current is basically part of armature current i.e.,

\[ i_s = i_a \quad \text{when } T_1 D_1 \text{ conducts} \]
\[ i_s = -i_a \quad \text{when } T_2 D_2 \text{ conducts} \]

and

\[ i_s = 0 \quad \text{when freewheeling diode conducts} \]

The waveform of Fig. 6.4.3 (e) shows the freewheeling diode current.

Mathematical analysis

We know that speed of the separately excited DC motor is given as,

\[ \omega = \frac{V_a - I_a R_a}{K_a \phi_f} \]

... (6.4.1)
Fig. 6.4.3 Waveforms of 1ϕ semiconverter based separately excited DC motor drive

Here $V_n$ is the output voltage of 1ϕ semiconverter. It is given as $\frac{V_m}{\pi}(1 + \cos \alpha)$. Hence above equation will be,

$$\omega = \frac{V_m (1 + \cos \alpha) - I_a R_a}{\pi K_a \phi_f}$$

... (6.4.2)

$$= \frac{V_m (1 + \cos \alpha)}{\pi K_a \phi_f} - \frac{I_a R_a}{K_a \phi_f}$$

This equation relates the speed with firing angle of semiconverter.
6.4.2.2 Discontinuous Load Current Mode

The motor current can be discontinuous if armature inductance is small. Fig. 6.4.4 shows the waveforms of discontinuous current mode.

Observe the armature current waveform of Fig. 6.4.4 (c). It is discontinuous. SCR $T_1$ is triggered at $\alpha$. Then current flows till $\beta$ beyond $\pi$. At $\beta$, $i_a = 0$. The next SCR $T_2$ is triggered at $\pi+\alpha$. Therefore $i_a$ again starts increasing from zero. The armature current again goes to zero at $\pi+\beta$. In the waveform of Fig. 6.4.4 (e) observe that freewheeling diode conducts from $\alpha$ to $\beta$ and $2\pi$ to $\pi+\beta$. In the waveform of Fig. 6.4.4 (c) observe that armature current is zero. But the motor rotates due to inertia. Hence the back emf of $e_b = k_f \phi N$ is generated across the terminals of the motor. Therefore $e_b$ is shown in the armature voltage waveform of Fig. 6.4.4 (b). Observe that $e_b$ appears across the armature whenever current is zero. That is

Fig. 6.4.4 Waveforms of 1φ semiconverter drive for discontinuous mode
c_b appears at β to (π+α), (π+β) to (2π+α) and so on. The motor is open circuited when we observe c_b across the motor terminals.

Mathematical analysis

The average of output voltage of the semiconverter for discontinuous mode is given as,

\[ V_{o(av)} = \frac{V_m}{\pi} (\cos \alpha - \cos \beta) \]

This \( V_{o(av)} \) is armature voltage \( (V_a) \) of the motor. Hence from equation (6.4.1) we get,

\[ \omega = \frac{\frac{V_m}{\pi} (\cos \alpha - \cos \beta) - I_a R_a}{K_n \phi_f} \quad \ldots \quad (6.4.3) \]

Disadvantages of discontinuous current mode

(i) Speed regulation is poor.
(ii) Motor performance is deteriorated
(iii) Peak current of the motor increases.
(iv) Dynamic response is slow.
(v) Motor efficiency is reduced.

6.4.3 1φ Full Converter Based DC Motor Drive

Full converters are better suitable for DC motors. Fig. 6.4.5 shows the circuit diagram of such drive.

![Fig. 6.4.5 1φ full converter based DC drive](image-url)
Observe that the armature is supplied by 1φ fully controlled bridge. The field is supplied with uncontrolled rectifier. Hence field flux remains constant. The operation of this drive can also be described for continuous mode and discontinuous mode.

6.4.3.1 Continuous Load Current Mode

Fig. 6.4.6 shows the waveforms for continuous armature current, SCRs $T_1$ and $T_2$ conduct in a positive half cycle. Since the load is inductive, the SCRs continue to conduct till next pair $T_3 T_4$ is triggered at $\pi+\alpha$. Observe that armature voltage is negative from $\pi$ to $\pi+\alpha$. This is second quadrant operation and it takes place due to inductive load. The armature current is shown in Fig. 6.4.6 (c). It is continuous. The ripple in armature current depends upon motor armature inductance. Fig. 6.4.6 (d) shows the supply current $i_s$. It is given as,

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---

Fig. 6.4.6 Continuous mode waveforms of 1φ FCB drive
\[ i_s = \begin{cases} 
  i_a & \text{when } T_1, T_2 \text{ conduct} \\
  -i_a & \text{when } T_3, T_4 \text{ conduct}
\end{cases} \]

Mathematical analysis

We know that average output voltage of the fully controlled bridge for continuous output current is given as,

\[ V_{dc(\alpha)} = \frac{2V_m}{\pi} \cos \alpha \]

Putting this value of \( V_{dc(\alpha)} \) for \( V_n \) in equation (6.4.1) we get,

\[ \omega = \frac{2V_m}{\pi} \cos \alpha - I_R R_a \]

\[ K_a \phi_f \]

This equation gives the speed in terms of firing angle (\( \alpha \)).

6.4.3.2 Discontinuous Load Current Mode

Fig. 6.4.7 shows the waveforms of fully controlled bridge drive for discontinuous operation. Fig. 6.4.7 (c) shows the armature current waveform. (See Fig. 6.4.7 on next page)

The current is discontinuous from \( \beta \) to \( \pi + \alpha \). During this period the motor is open and no current flows. Since the motor is rotating due to inertia, a back emf of \( e_b = K_a \phi_f N \) is developed across its terminals. The back emf appears across the motor terminals when armature current is zero. This is shown in the waveform of Fig. 6.4.7 (b). The supply current is shown in Fig. 6.4.7 (d). The discontinuous mode has similar drawbacks as discussed for the semiconverter drive in last subsection.

Example 6.4.1: A separately excited DC motor is operated from a single phase semiconverter and has following parameters: \( R_a = 0.05 \Omega \), \( k_a \phi_f = 1 \text{N-m/A} \). The supply voltage is 230V/50Hz. Calculate the speed of the motor for a torque of 8 N-m and firing angle of 60°.

Solution:

Here \( k_a \phi_f = 1 \text{N-m/A} \)

\[ T = 8 \text{ N-m} \]

\[ R_a = 0.05 \Omega \]

\[ V_m = 230\sqrt{2} \text{ V} \]

For semiconverter, speed of the motor is given by equation 6.4.2 as,
Fig. 6.4.7 Discontinuous mode waveforms of 1Φ FCB based drive.

\[
\omega = \frac{V_m (1 + \cos \alpha) - I_a R_a}{K_a \phi_f}.
\]

We know that \( I_a = \frac{T}{K_a \phi_f} \). Hence above equation will be,

\[
\omega = \frac{V_m (1 + \cos \alpha) - T R_a}{K_a \phi_f}.
\]

\[
230 \sqrt{2} (1 + \cos 60^\circ) = 230 \times \frac{1}{2} = 115\sqrt{2}
\]

\[
= \left( \frac{V_m}{K_a \phi_f} \right) \left( \frac{T R_a}{K_a \phi_f} \right)
\]

\[
= \frac{V_m (1 + \cos \alpha) - 115}{K_a \phi_f}
\]

\[
= \frac{155 \text{ rad/sec}}{1}
\]
VARIABLE-FREQUENCY CONVERTER CLASSIFICATIONS

Based on the discussion in the previous section, the variable-frequency converters, which act as an interface between the utility power system and the induction motor, must satisfy the following basic requirements:

1. Ability to adjust the frequency according to the desired output speed
2. Ability to adjust the output voltage so as to maintain a constant air gap flux in the constant-torque region
3. Ability to supply a rated current on a continuous basis at any frequency

Except for a few special cases of very high power applications where cycloconverters are used (these are briefly discussed in Chapter 15), variable-frequency drives employ inverters with a dc input, as discussed in Chapter 8. Figure 14-17 illustrates the basic concept where the utility input is converted into dc by means of either a controlled or an uncontrolled rectifier and then inverted to provide three phase voltages and currents to the motor, adjustable in magnitude and frequency. These converters can be classified based on the type of rectifier and inverter used in Fig. 14-17:

- Pulse-width-modulated voltage source inverter (PWM-VSI) with a diode rectifier
- Square-wave voltage source inverter (square-wave VSI) with a thyristor rectifier
- Current source inverter (CSI) with a thyristor rectifier

As the names imply, the basic difference between the VSI and the CSI is the following: In the VSI, the dc input appears as a dc voltage source (ideally with no internal impedance) to the inverter. On the other hand, in the CSI, the dc input appears as a dc current source (ideally with the internal impedance approaching infinity) to the inverter.

Figure 14-18a shows the schematic of a PWM-VSI with a diode rectifier. In the square-wave VSI of Fig. 14-18b, a controlled rectifier is used at the front end and the inverter operates in a square-wave mode (also called the six-step). The line voltage may be single phase or three phase. In both VSI controllers, a large dc bus capacitor is used to make the input to the inverter appear as a voltage source with a very small internal impedance at the inverter switching frequency.

From the schematic of VSI converters shown in Figs. 14-18a and 14-18b, it is recognized that the switch-mode, dc-to-ac VSIs have been discussed previously in Chapter 8 in both square-wave and PWM modes of operation. It should be noted that, in practice, only three-phase motors are controlled by means of variable frequency. Therefore, only the dc-to-three-phase-ac inverters are applicable here. Also, the controlled and uncontrolled (diode) rectification of single-phase and three-phase ac inputs to dc has been discussed in detail in Chapters 5 and 6. Therefore, the main emphasis in this chapter will be on the interaction of VSIs with induction motor type of loads.

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Figure 14-17 Variable-frequency converter.

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Converters, which must satisfy speed and gap flux in the frequency. Cycloconverters drives employ the basic controlled or an controlled based on a diode rectifier thyristor rectifier. The CSI is the full diode with no internal rectifier to the inverter. In the front end and the line voltage may capacitor is used very small internal and 14-18b, it is previously in Chapter be noted that, in frequency. Therefore, the controlled and puts to dc has been in this chapter will

Figure 14-18 Classification of variable-frequency converters: (c) PWM-VSI with a diode rectifier; (b) square-wave VSI with a controlled rectifier; (c) CSI with a controlled rectifier.

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Figure 14-18c shows the schematic of a CSI drive where a line-voltage-commutated controlled converter (discussed in Chapter 5) is used at the front end. Because of a large inductor in the dc link, the input to the inverter appears as a dc current source. The inverter utilizes thyristors, diodes, and capacitors for forced commutation.

14-7 VARIABLE-FREQUENCY PWM-VSI DRIVES

Figure 14-19a shows the schematic of a PWM-VSI drive, assuming a three-phase utility input. As a brief review of what has already been covered in Chapter 8, the PWM inverter controls both the frequency and the magnitude of the voltage output. Therefore, at the input, an uncontrolled diode bridge rectifier is generally used. One possible method of generating the inverter switch control signals is by comparing three sinusoidal control voltages (at the desired output frequency and proportional to the output voltage magnitude) with a triangular waveform at a selected switching frequency, as shown in Fig. 14-19b.

As discussed in Chapter 5, in a PWM inverter, the harmonics in the output voltage appear as sidebands of the switching frequency and its multiples. Therefore, a high