3 Multiple Connections of SCRs

3.1 SERIES-PARALLEL OPERATION OF SCRs

Silicon-controlled rectifiers are now available with a voltage rating upto 10 kV and a current rating upto 1200 A. In many power-control applications, the required voltage and current ratings are lower than these maximum limits. Therefore, even though it may be possible to obtain a single SCR of proper voltage and current ratings, the designer on many occasions is forced to use lower-rated SCRs for reasons of economy and availability. In such a situation, the lower-rated SCRs have to be connected in series and parallel combinations to suit the voltage and current requirements of the circuit for a particular application. The series and parallel combinations are also often used when it is required to control power in low-voltage high-current circuits or high-voltage low-current circuits because an SCR of suitable voltage and current ratings may not be available.

In this chapter, we will consider the problems associated with series and parallel connections of SCRs, and discuss how these may be overcome so that the SCRs can be utilised to the fullest advantage by improving their string efficiency. This efficiency is defined as the ratio of the total current/voltage rating of the whole string to the individual current/voltage rating of the SCR, multiplied by the number of units in parallel/series in the string. In practice, this ratio is always less than one. For achieving the best efficiency, the characteristics of all SCRs used in the string must be the same. Since it is impossible to get SCRs of identical characteristics, these have to be matched as far as possible. Small deviations in characteristics lead to unequal voltage/current distribution in the units connected in series/parallel. Figure 3.1a shows how voltage is shared by two SCRs of the same voltage rating when connected in series. Since they have to carry the same current, the SCR with the higher blocking resistance R_D ($=dV/dI_D$) will share a larger portion of the applied voltage. Similarly, Fig. 3.1b shows the dynamic characteristics of two SCRs connected in parallel. Since the voltage drop across the two devices must be the same, the SCR with the lower R_T ($=dV/dI_T$) will carry more current. The unequal sharing of voltage/current can be corrected by external equalising circuits.

3.2 SERIES OPERATION

When it is desired to increase the voltage rating of a string, the SCRs can be connected in series as shown in Fig. 3.2a. As already mentioned, these SCRs have to be properly matched as far as possible. Let us suppose that there are some differences in the individual characteristics of the SCRs.

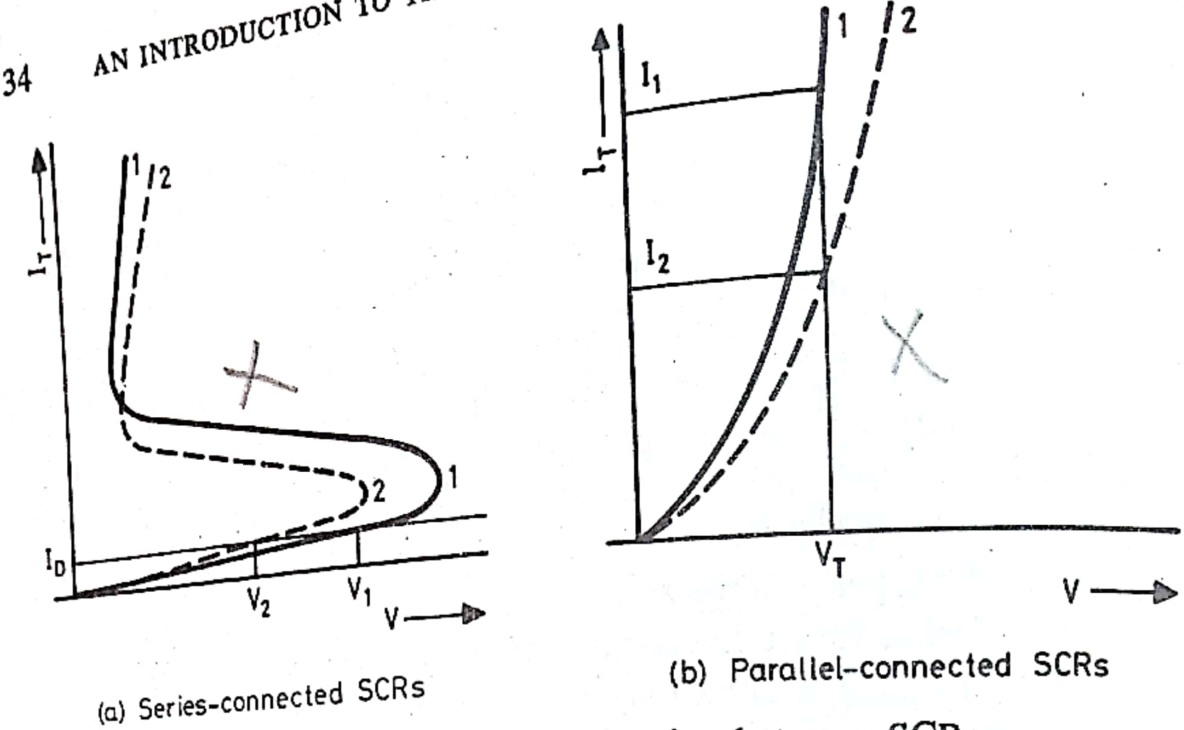


Fig. 3.1 Voltage/current sharing between SCRs.

For example, assume that SCR1 has a high turn-on time. Further, SCR2 has the lowest forward and reverse blocking currents, and the recovery time t_{rr} of SCR3 is small. Figure 3.2b shows the voltage distribution across

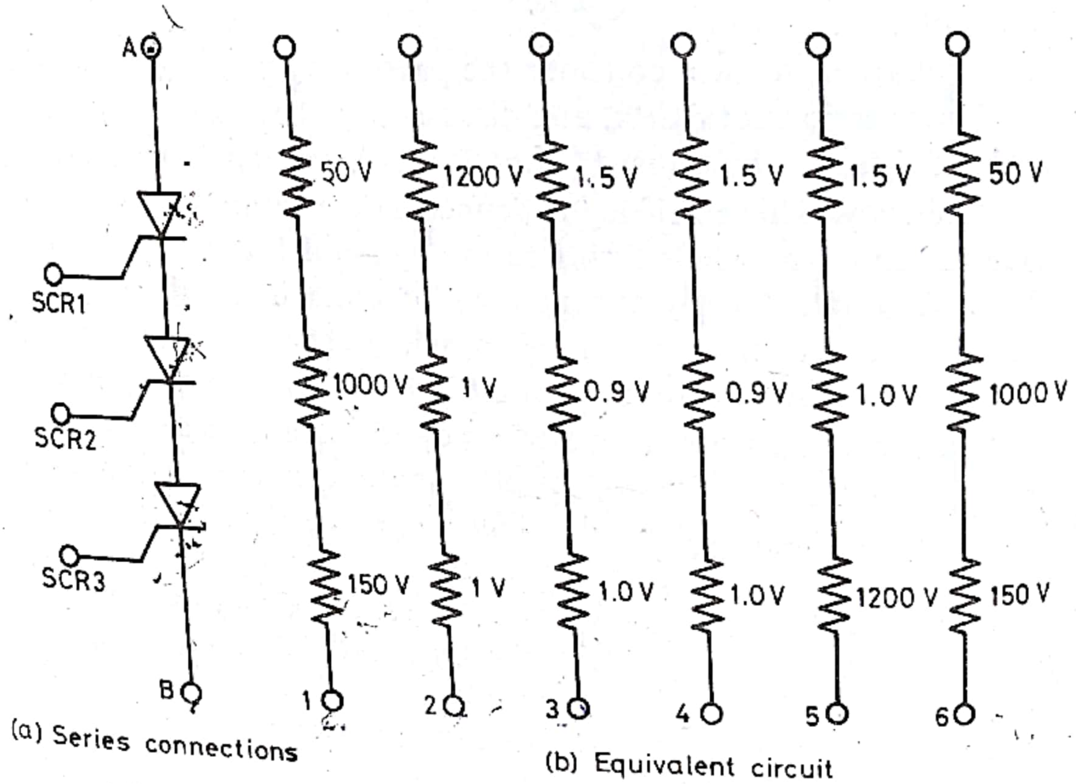


Fig. 3.2 Voltage distribution in series-connected SCRs.

the string of SCRs for six different situations, namely, (1) all SCRs in the forward blocking state, (2) immediately after turn-on, (3) all SCRs conducting, (4) reverse voltage applied (all SCRs conducting in the reverse direction), (5) only SCR3 recovered, and (6) all SCRs in the reverse blocking state (all SCRs recovered). For these cases, the forward and reverse voltages applied are taken to be 1200 V.

It can be observed that even though each SCR is identically rated, the voltage distribution is not uniform because of some differences in the characteristics, and that the SCR with the lowest voltage will break down, leading to the breakdown of the whole string. Therefore, some external compensating circuit is required to produce a uniform voltage distribution under all conditions of operation. For conditions (3) and (4), in which all the SCRs are conducting, a small unequal voltage drop will not affect the performance of the series circuit. For conditions (1) and (6), when the string is blocking the forward and reverse voltages, respectively, the voltage distribution can be made more uniform by connecting a shunt resistor across each SCR as shown in Fig. 3.3. This is called the *static equalising*

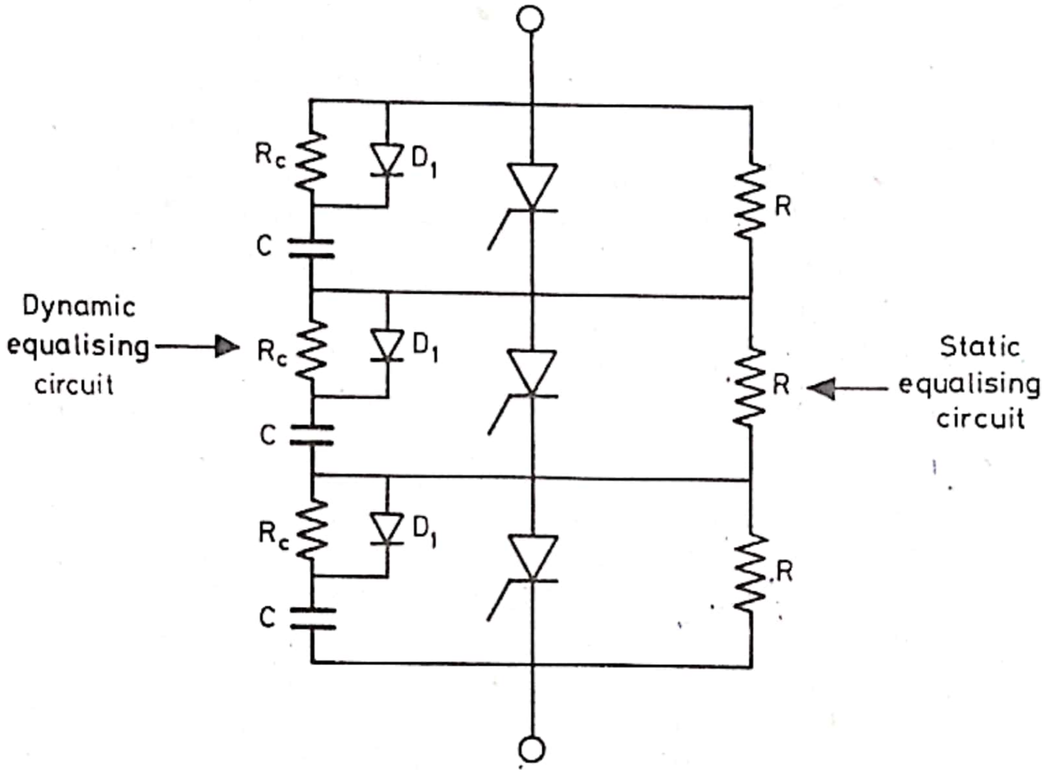


Fig. 3.3 Equalising circuits for series operation.

circuit. The value of the shunt resistance R is given by

$$R = \frac{nV_D - V_S}{(n-1)I_B}, \tag{3.1}$$

where n is the number of SCRs in series, V_D is the voltage rating of each SCR, V_S is the total voltage across the string, and I_B is the maximum blocking current of the SCRs at the rated voltage. This shunt resistance reduces the effect of the different blocking resistances of the SCRs, and thus produces a more uniform voltage distribution.

During the turn-on and turn-off periods, corresponding to operating conditions (2) and (5), respectively, because of the transient nature of the voltage and current, a simple resistance divider will not equalise the voltage. Here, the capacitance of the reverse-biased junction controls the voltage distribution, and therefore requires external shunt capacitance, as shown in Fig. 3.3, to provide uniform voltage distribution, and also to improve the dv/dt rating of the device. This is called the *dynamic equalis-*

ing circuit. During the period for which any of the devices is in the blocking state, the corresponding capacitor will get charged to the voltage across that unit. When the device is turned on, there will be a heavy discharge current from the capacitors. To limit this discharge current, a small resistor R_c is used in series with the capacitor. Diode D_1 will cut off this resistor during the charging time of the capacitor when forward voltage is applied to the string. The value of C is obtained from

$$C = \frac{(n-1)\Delta Q}{nV_D - V_S}, \tag{3.2}$$

where ΔQ is the maximum difference in the recovery charge of the SCRs in the string. The value of ΔQ for various SCRs can be obtained from handbooks or from the application notes of the manufacturers. The choice of R_c depends on the permissible peak repetitive current through the SCR. During the turn-off period, when one SCR in the string has completely recovered, the shunt resistor R will provide an alternate path for the reverse current of the other SCRs to flow through. This facilitates the turn-off of the whole string. The derivations for Eqs. (3.1) and (3.2) are given in Example 3.5.

3.2.1 TRIGGERING OF SERIES-CONNECTED SCRs

In spite of voltage equalising circuits, the sharing of voltage will be non-uniform if all the SCRs are not triggered simultaneously. For proper functioning of the series string, a gate signal of sufficient amplitude should be applied to all the gates of the SCRs at the same time. The small differences in turn-on times are properly corrected by the dynamic equalising circuit. Figure 3.4a shows one method of firing the string. The main

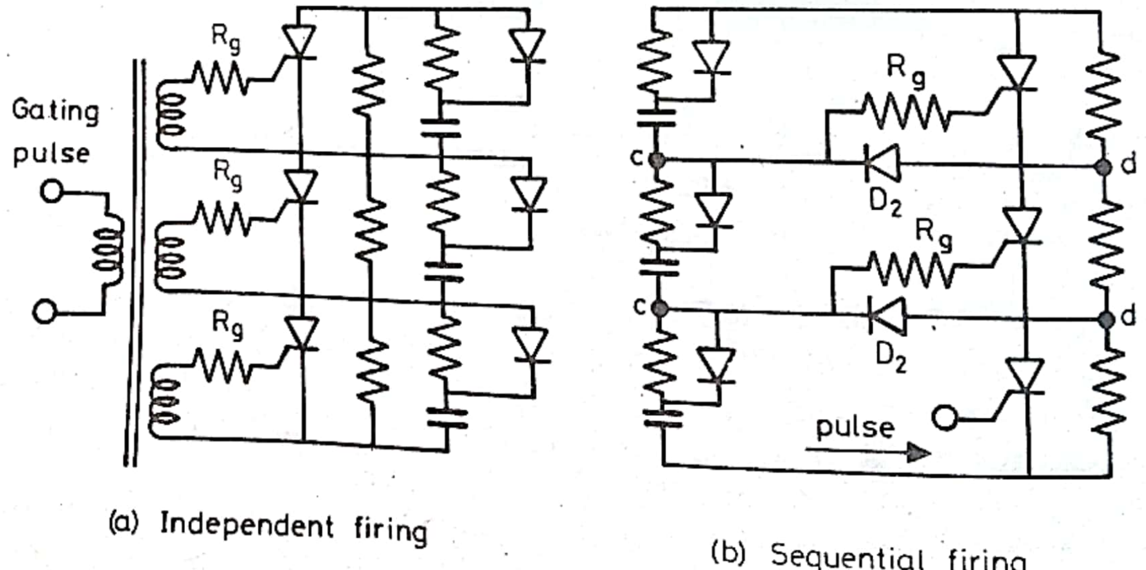


Fig. 3.4 Circuits for triggering series-connected SCRs.

triggering pulse is applied to the primary of the transformer. Each of the secondary windings is connected to individual gates of respective SCRs in the string as shown in the figure. To equalise the gate current in each SCR, a resistor R_g is connected in series with the secondary winding for swamping out any difference in the gate-to-cathode impedance of individual units.

Figure 3.4b shows another method of triggering the series string. Here, only one SCR at the bottom of the string is turned on by the external gate pulse. The discharge current of the shunt capacitor, through the SCR which is turned on, will fire the next SCR in the string. This process takes place rapidly, and all the SCRs are turned on in a very short time. Since the SCRs are turned on in sequence, the topmost SCR in the string will experience an increasing forward voltage. This sequential firing of series-connected SCRs is used for generating impulse voltages. The value of capacitance C, when required to provide uniform voltage distribution and also the necessary trigger current for the SCR, is given [in microfarads (μF)] by

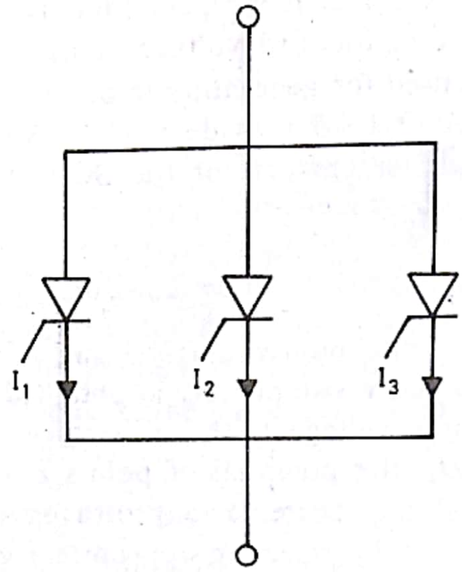
$$C = \frac{10}{R_g + V_{GT}/I_{GT}}, \tag{3.3}$$

where V_{GT} and I_{GT} are the maximum gate-triggering voltage and current, respectively. The gate source resistance R_g is obtained as explained in Section 2.8. The gate source voltage is the off-state voltage across each SCR. Because of diode D_2, the potential of points c, d will not be the same. The resulting circulating currents may turn on the SCRs. These currents must be minimised by selecting appropriate equalising circuit parameters so that the string is turned on only when a gate signal is applied to the bottom SCR.

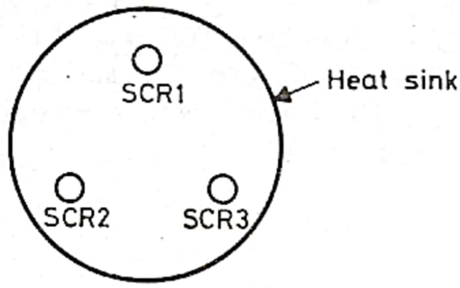
Another convenient method of firing a string of SCRs employs optical triggering. In this case, a light-activated SCR (LASCR) is connected between the gate and capacitor C through a suitable resistance. One LASCR is used for each stage. The LASCR is fired by photon bombardment and capacitor C discharges through the gate and turns on the SCR.

3.3 PARALLEL OPERATION

Silicon-controlled rectifiers are connected in parallel to improve the current rating. Due to the unequal dynamic resistance R_T of the SCRs, the sharing of current will not be equal. This problem leads to what is known as *thermal runaway*. For example, if one of the SCRs in a parallel unit carries more current than the other members, its internal power dissipation will be more, thereby raising the junction temperature and decreasing the dynamic resistance of the SCR. This, in turn, will increase the current shared by this SCR and the process becomes repetitive, the cumulative increase in current results in permanent damage to the SCR, followed by the burning out of other SCRs, one by one. Therefore, one important precaution to be observed when SCRs are operated in parallel is that, as far as possible, all units must operate at the same temperature. This can be done by having a common heat sink. Unequal sharing of current is also the result of the inductive effect on current-carrying conductors. If three SCRs are arranged as shown in Fig. 3.5a, the middle one will have more flux linkages and, therefore, more inductance. In AC circuits, the reactance drop for the central limb will be higher, and so less current will flow through it. The outer two parallel units carry more current. This uneven distribution of current is corrected by having a symmetrical arrangement of SCRs as shown in Fig. 3.5b, where the SCRs are arranged in a circular configuration on the same heat sink.



(a) Flat arrangement



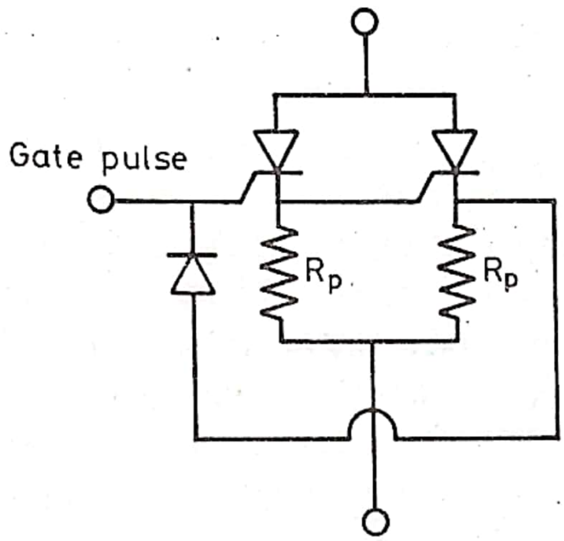
(b) Symmetrical arrangement

Fig. 3.5 Configurations of SCRs for parallel operation.

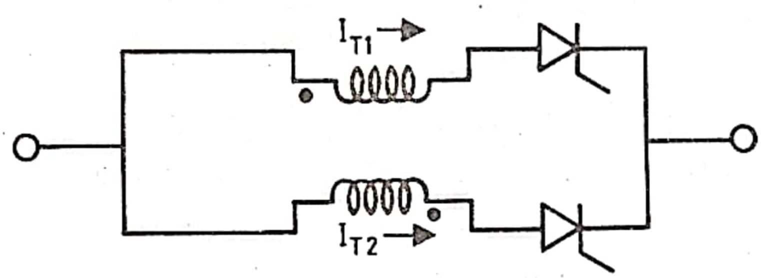
In DC circuits, the difference in the value of dynamic resistance R_T is compensated by connecting a resistor R_p in series with each SCR, as shown in Fig. 3.6a. If the two SCRs are of identical rating, then the two external series resistors R_{p1} and R_{p2} are chosen such that the total voltage drops are equal, that is, $R_{p1} + R_{T1} = R_{p2} + R_{T2}$, where R_{T1} and R_{T2} are the corresponding dynamic resistances of the two devices at the rated current I_T. If two SCRs of different forward current ratings I_{T1} and I_{T2} are to be operated in parallel, then the same resistance R_p can be used for both units to ensure proper current sharing by the SCRs. Let V_{T1} and V_{T2} be the respective voltage drops across the two SCRs for forward currents I_{T1} and I_{T2}. Since the units are in parallel, their anode-to-cathode voltage drops must be the same. Therefore,

$$V_{T1} + I_{T1}(R_p + R_{T1}) = V_{T2} + I_{T2}(R_p + R_{T2}). \tag{3.4}$$

The value of R_p can be calculated from Eq. (3.4). The disadvantage of this type of compensation is that there is considerable loss of power due to series resistance.



(a) Resistance compensation



(b) Inductive compensation

Fig. 3.6 Equalising circuits for parallel-connected SCRs.

In AC circuits, the sharing of current can be made uniform by the magnetic coupling of parallel paths as shown in Fig. 3.6b. If currents I_{T1} and I_{T2} are equal, then the voltage drop in the transformer will be zero because of the mutual cancellation of flux linkages in the coils. If for any reason the currents are unequal, the transformer will produce an induced voltage in the windings which will reduce the difference between I_{T1} and I_{T2}.

6 AC Power Control

6.1 PHASE CONTROL

In Chapter 2, we considered various methods for triggering SCRs. Of these, the most efficient method for power modulation with thyristors uses gate control. In AC circuits, the SCR can be turned on by the gate at any angle α with respect to the applied voltage. This angle α is the *firing angle.* Power control is obtained by varying the firing angle, and this is known as *phase control.*

Figure 6.1a shows a simple half-wave circuit. The principle of phase control for an inductive load can be explained by this circuit. Let the SCR be fired at an angle α. The load current, load voltage, and supply voltage waveforms are shown in Fig. 6.1b. The SCR will turn off by natural commutation when the current becomes zero. Angle β is known as the *conduction angle.* By varying the firing angle α, the RMS value of the load voltage can be changed. The firing circuits discussed in Chapter 2 can be used to control the firing angle. Current $i(t)$ in the circuit shown in Fig. 6.1 is given by

$$i(t) = \frac{E_m}{\sqrt{(R^2 + \omega^2 L^2)}} \sin(\omega t + \alpha - \phi) + \frac{E_m}{\sqrt{(R^2 + \omega^2 L^2)}} \sin(\phi - \alpha) e^{-Rt/L} \quad \text{for } 0 \leqslant t \leqslant \beta/\omega, \tag{6.1}$$

where ϕ is $\tan^{-1}(\omega L/R)$ and E_m is the amplitude of the input voltage. The average value of the load current can be obtained from the equation for $i(t)$. The power consumed by the load decreases as angle α is increased. The reactive power input from the supply increases with the firing angle. The load current waveform can be improved by connecting a free-wheeling diode D_1 as shown by the dashed line in Fig. 6.1a. With the diode, SCR1 will be turned off as soon as the input voltage polarity reverses. After that, the load current will free-wheel through the diode and a reverse voltage will appear across the SCR. Figure 6.1c shows the load current and load voltage waveforms with a free-wheeling diode. There are two modes of operation for this circuit. In the first mode, diode D_1 will be reverse-biased and SCR1 will conduct. This mode will exist from the instant of firing to the time when the voltage polarity reverses. The duration is given by $(\pi - \alpha)/\omega$. The current waveform is

$$i(t) = \frac{E_m}{\sqrt{(R^2 + \omega^2 L^2)}} \sin(\omega t + \alpha - \phi) + \left\{\frac{E_m}{\sqrt{(R^2 + \omega^2 L^2)}} \sin(\phi - \alpha) + i_0\right\} e^{-Rt/L} \quad \text{for } 0 \leqslant t \leqslant (\pi - \alpha)/\omega, \tag{6.2}$$

where i_0 is the current in the load when the SCR is fired. Let the current at the end of mode 1 $[t = (\pi - \alpha)/\omega]$ be I_o. During mode 2, SCR1 will

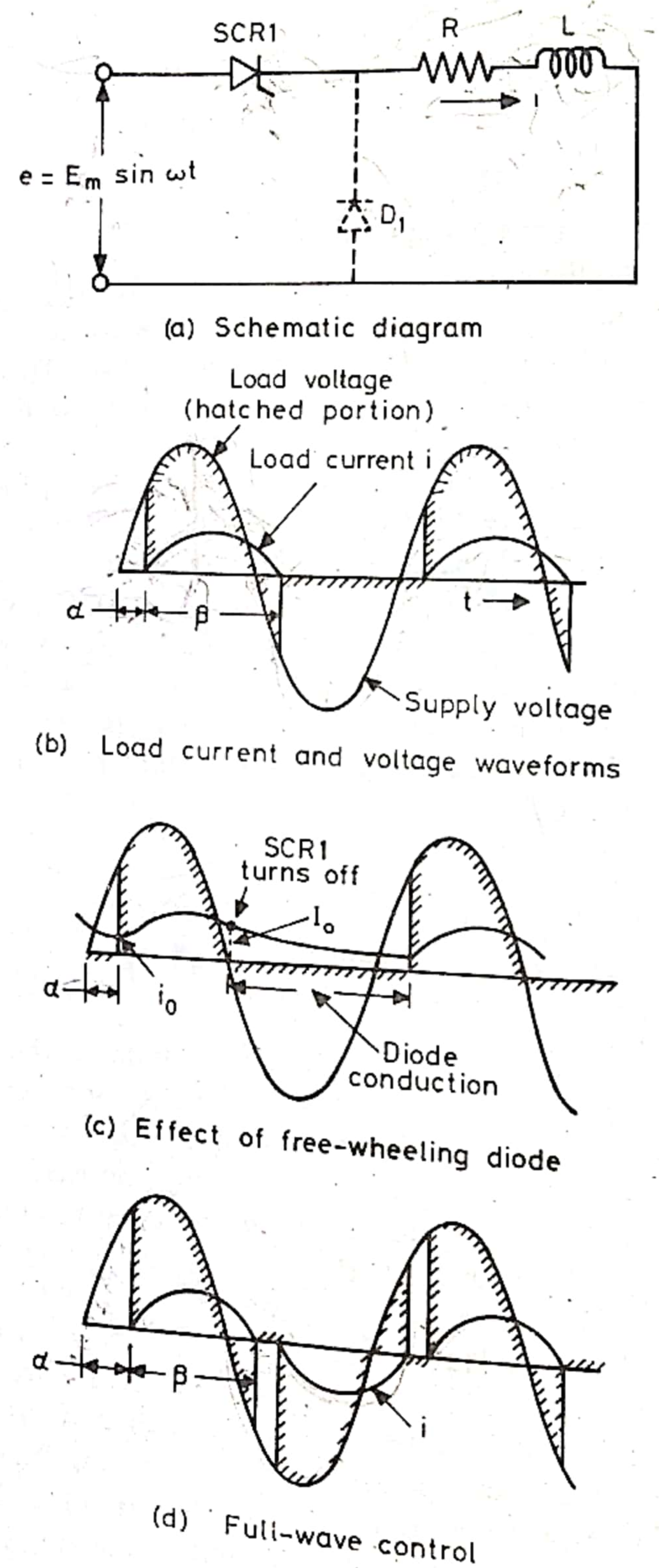


Fig. 6.1 Phase control.

be reverse-biased and diode D_1 will conduct. It is assumed that the load is sufficiently inductive to maintain the current in the load circuit until the next instant of firing. Current $i(t)$ in this period is given by

$$i(t) = I_0 e^{-Rt/L} \quad \text{for } (\pi - \alpha)/\omega \leqslant t \leqslant 2\pi/\omega. \tag{6.3}$$

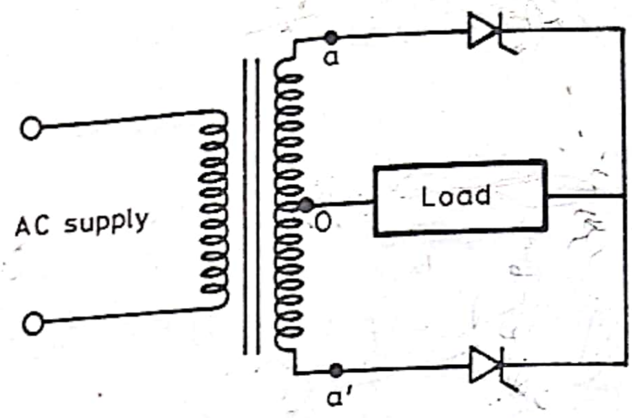
In the steady state, the current at the end of mode 2 must be equal to i_0. Comparing the load current waveforms in Figs. 6.1b and 6.1c, it will be observed that, for the same firing angle, the load power consumption is more with a free-wheeling diode. The power flow from the input takes place only during mode 1. Therefore, the ratio of reactive power flow from the input to the total power consumed in the load is less for the phase-control circuit with a free-wheeling diode. In other words, the free-wheeling diode improves the input power factor. This is because the inductive energy of the load is dissipated in the load resistance R during mode 2 instead of returning to the input.

In the half-wave circuits just discussed, the input current has a large DC component since the current flow is unidirectional. This asymmetrical current produces magnetic saturation in the input transformers, if any are used. However, this problem is remedied in three-phase half-wave controlled circuits by the special winding connections on the secondary side of the input transformer. The input power factor can also be improved by connecting the primary windings in delta.

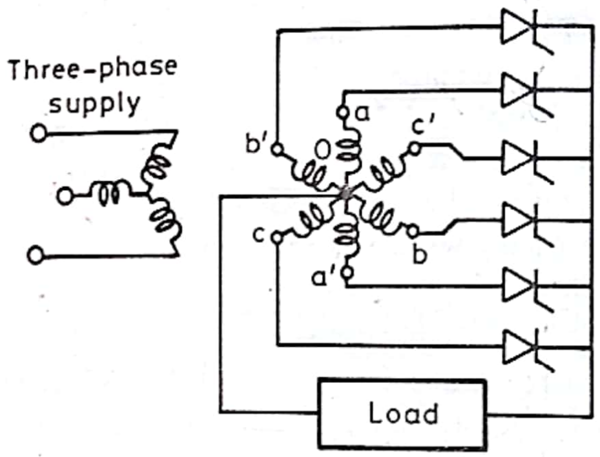
6.2 FULL-WAVE CONTROL CIRCUIT

Full-wave power control is obtained by replacing SCR1 in Fig. 6.1a by a triac or two SCRs in antiparallel. A free-wheeling diode cannot be used for such a circuit. Symmetrical firing is used in each half-cycle. The resulting current waveform is shown in Fig. 6.1d. Any one of the SCRs can be fired only after the other conducting SCR is turned off due to natural commutation. This limits the minimum firing angle α to the impedance angle ϕ of the load circuit at the input frequency. Phase control is also used for obtaining a variable DC voltage. The half-wave control circuit shown in Fig. 6.1a can be employed for this purpose. *LC* filters are utilised for reducing the ripple in the output and the load is connected on the DC side. The filter size can be reduced by using full-wave control circuits (with polyphase input) because these circuits produce higher ripple frequency in the output as compared with half-wave control circuits. There are two basic configurations for full-wave control circuits. One configuration requires an input transformer with two identical windings on the secondary side for each phase. The two windings have a common terminal. This is known as the *midpoint configuration*. Single-phase circuits require two SCRs (M-2 connection) and three-phase circuits need six SCRs (M-6 connection) as shown in Figs. 6.2a and 6.2b. These are known also as two-pulse and six-pulse converters. The number of pulses is equal to the order of the lowest harmonic in the output. Variable DC voltage is obtained by controlling the firing angle. Hence, these circuits are also referred to as *controlled rectifiers* or *converters*.

The second configuration, called the *bridge circuit*, is shown in Fig. 6.3. Here, no input transformer is required. The single-phase circuit (Fig. 6.3a)



(a) M-2 connection



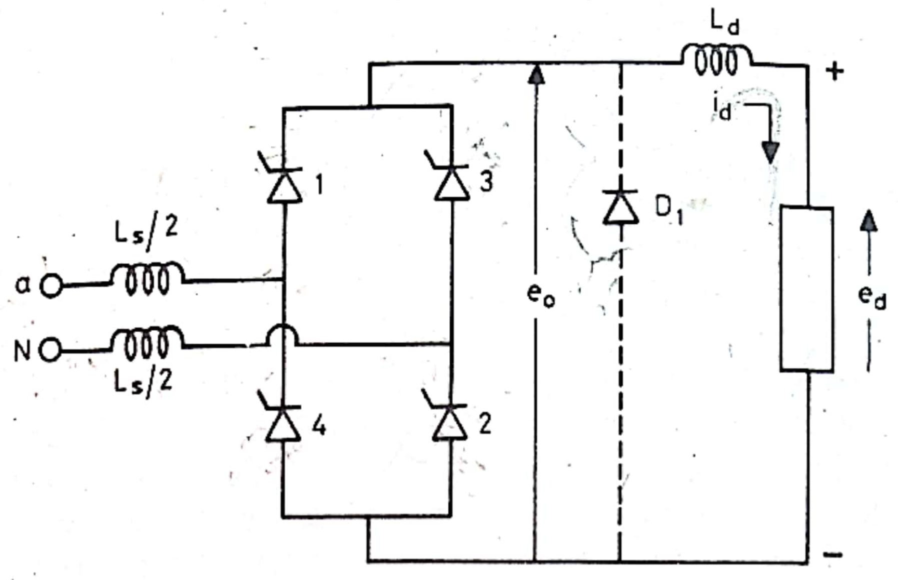
(b) M-6 connection

Fig. 6.2 Full-wave control circuits.

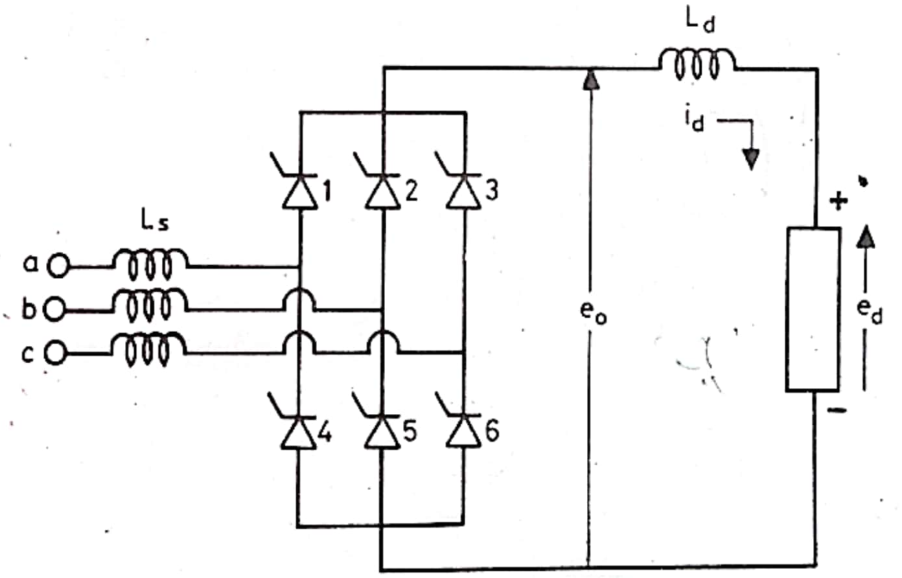
is known as the B-2 connection and the three-phase circuit (Fig. 6.3b) is known as the B-6 connection. The numerals in these notations correspond to the number of pulses in the output during one period of the input wave. For a given voltage rating of the SCRs, the load voltage for the M-2 connection is one-half that for the B-2 connection. The volt-ampere rating of the transformer in Fig. 6.2a is twice that of the load. Therefore, the bridge configuration is preferable unless one of the terminals on the DC side has to be grounded.

6.2.1 ANALYSIS OF A BRIDGE CIRCUIT

The single-phase circuit of Fig. 6.3a is considered for the analysis. Inductance L_d is used in the DC circuit to reduce the ripple. A large value of L_d will result in a continuous steady current in the load. A small L_d will produce a discontinuous load current for large firing angles. The source is assumed to have an internal inductance L_s. Let SCRs 1 and 2 be fired at an angle α in the positive half-cycle. The direction of load current i_d is as shown in the figure. The current waveforms for the two extreme values of L_d are given in Figs. 6.4a and 6.4b. The effect of source inductance is neglected here. It can be seen that for this circuit if the DC-side negative



(a) Single-phase circuit



(b) Three-phase circuit

Fig. 6.3 Bridge configurations.

terminal is grounded (connected to the supply neutral), SCRs 2 and 3 will not fire and there will be a short-circuit on the supply during the negative half-cycle when SCR4 is triggered. The average DC voltage e_d is $(2E_m/\pi)\cos\alpha$.

In Fig. 6.4a, since the smoothing inductor is very large, current i_d in the steady state will be pure DC. Therefore, even when the input voltage polarity is reversed, the current will continue to flow through SCRs 1 and 2 till the other pair, SCRs 3 and 4, is fired symmetrically at an angle α in the negative half-cycle. Since the polarity of the input voltage is already reversed, the firing of SCRs 3 and 4 will reverse-bias SCRs 1 and 2, and turn them off. The load current will then shift from the pair of SCRs 1 and 2 to the pair of SCRs 3 and 4. This is referred to as *class F type forced commutation* (see Section 8.2), or *line commutation*. Current i_d will maintain the same direction of flow in the load. However, the current on the input side will flow in the reverse direction when SCRs 3 and 4 conduct, that is, the input current i_L will be a rectangular AC wave. The output voltage e_o is shown by the hatched lines in Fig. 6.4a; the dashed line corresponds to the average value e of this output voltage. When L_d is zero and source inductance is neglected, current i_d will go to zero at the end of every half-cycle (Fig. 6.4b). Thus, no current will flow through the load from the end of the half-cycle to the instant when the other pair of SCRs is fired. When L_d is small, current i_d will still go to zero before the other pair of SCRs is fired if the firing angle α is sufficiently large. Here also, the conducting pair of SCRs will turn off due to natural commutation. The load voltage e is approximately equal to e_o. An expression for the minimum value of L_d to provide continuous current is derived in Chapter 7. For large values of L_d, voltage e_d is steady since i_d is DC. This will be the average value of voltage e_o (shown by the hatched portion in Fig. 6.4a). The firing angle α for the SCRs can be changed from zero to π. During this period, the potential of the incoming SCRs will be more than that of the conducting SCRs, and proper commutation can take place.

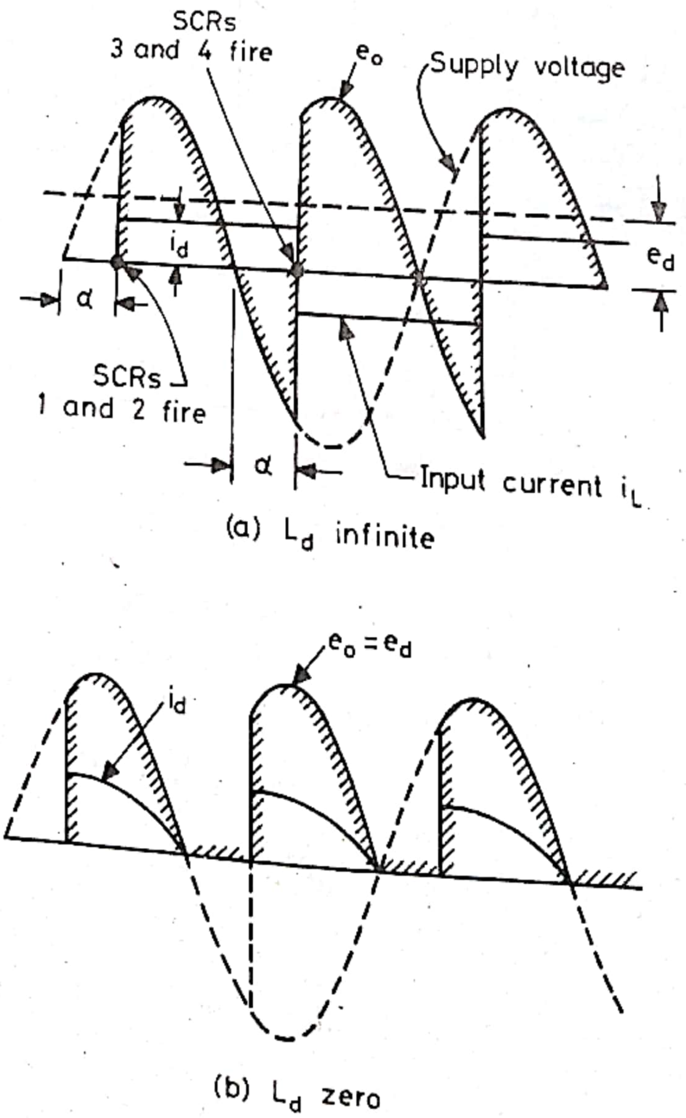


Fig. 6.4 Current and voltage waveforms for bridge circuits (*cont.*).

Heyden, London, 1972. ... converters,

6.4 DUAL CONVERTERS

As the name indicates, a dual converter consists of two converters, both either fully-controlled or half-controlled, connected to the same load. The purpose of a dual converter is to provide a reversible DC voltage to the load. It is needed for DC motor drives where speed reversal is required. Figure 6.8 shows the schematic arrangement of a dual converter. The two modes of its operation are the noncirculating-current mode and the circulating-current mode. In the former, only one of the bridges is triggered.

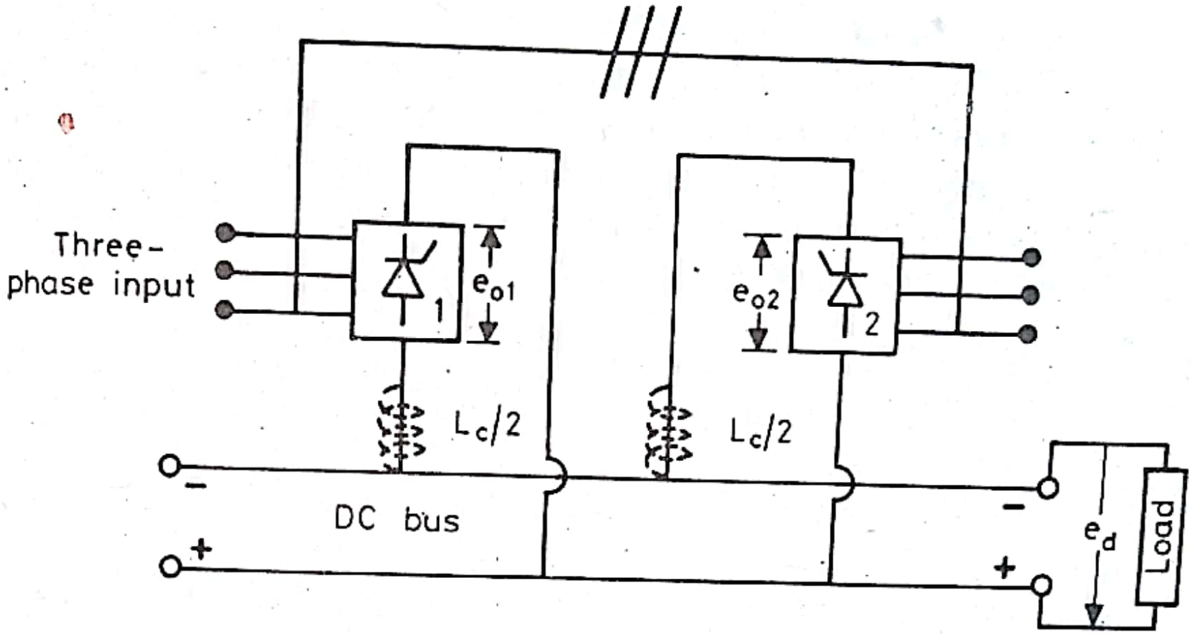


Fig. 6.8 Dual converter.

When reversal of output voltage is required, the firing pulses for the conducting bridge are stopped and the second bridge is gated. Since the conducting SCRs in the first bridge will turn off only when the current goes to zero, a small dead time must be allowed before the second bridge is gated. Otherwise, the AC input will be shorted through the two bridges. For this, a current sensor is required which ensures that all the SCRs in bridge 1 will be turned off before the firing pulses are applied to SCRs of the second bridge. In the circulating-current mode, both bridges are gated simultaneously, one operating in the rectifying mode and the other in the inverting mode to avoid short-circuits. This scheme requires fully-controlled bridges. For the polarity shown in Fig. 6.8, bridge 1 is a rectifier and bridge 2 is an inverter. The internal voltage of the rectifier is higher and that of the inverter is lower than the output voltage. If the firing angle for bridge 1 is α, then its internal voltage $(3\sqrt{3}/\pi)E_m \cos\alpha$ (where E_m is the peak value of the input phase-to-neutral voltage) is made slightly more than the DC output voltage e_d; the firing angle for bridge 2 is made slightly less than $(\pi - \alpha)$ so that its internal voltage is lower than e_d. If the reversal of output voltage polarity is required, then the firing angles of the two bridges are changed simultaneously such that bridge 2 will operate as a rectifier and bridge 1 as an inverter. The polarity of the output voltage will be the same as that of the rectifier. The internal voltage of the inverter must be close to but smaller than the DC output voltage. This is to ensure that the circulating currents between the two bridges are minimised. The main advantage of the circulating current scheme is the rapidity with which the phase reversal of the output current can be obtained. However, this scheme will produce a continuous flow of circulating current between the two bridges, resulting in increased power losses. A similar scheme is also used for cycloconverters for AC-to-AC conversion. This method of frequency conversion and firing angle control will be explained in Chapter 7.

To reduce the circulating current, it is necessary to include inductance L_c (shown by the dashed lines in Fig. 6.8) in the circulating-current path. The load is connected to the centre tap of the coil. Even though the firing angles of the two bridges are adjusted in a manner such that their average output voltages are almost equal, there will be a difference in the instantaneous values of these output voltages when one of the bridges is operating as a rectifier and the other as an inverter. For a two-pulse converter, the output voltage e_o will be

$$e_o = \frac{2E_m}{\pi}[\cos\alpha + (\frac{\cos 3\alpha}{3} - \cos\alpha)\cos 2\omega t + (\frac{\sin 3\alpha}{3} - \sin\alpha)\sin 2\omega t + \ldots], \tag{6.8}$$

where E_m is the peak value of the AC input voltage and α is the firing angle. If α and $(\pi - \alpha)$ are the firing angles for the two converters, then the sum of the two voltages $(e_{o1} + e_{o2})$ will be twice the sine component of the second harmonic (neglecting all other higher harmonics) which drives the circulating current. This current will have a maximum peak value when $\alpha = \pi/2$, and the corresponding average load voltage and current will be ... suming that the maximum peak circulating current is ...

$\times 0.56 = 10.08$.

6.6 APPLICATION TO SPEED CONTROL OF MOTORS

Phase control can be very conveniently used for the speed control of AC and DC motors; this is achieved by applying a variable voltage to the motor. As the speed of synchronous motors does not change when the input voltage is varied, this method is useful only for commutator or induction motors. For AC motors, full-wave phase-control circuits are required. Figure 6.11 shows the schematic arrangement for the speed control of single-phase and three-phase induction motors. By varying the firing angle, the RMS input voltage can be changed. In the case of single-phase motors, an additional starting winding is required. Two SCRs connected in anti-parallel are preferred to one triac since the motor is an inductive load.

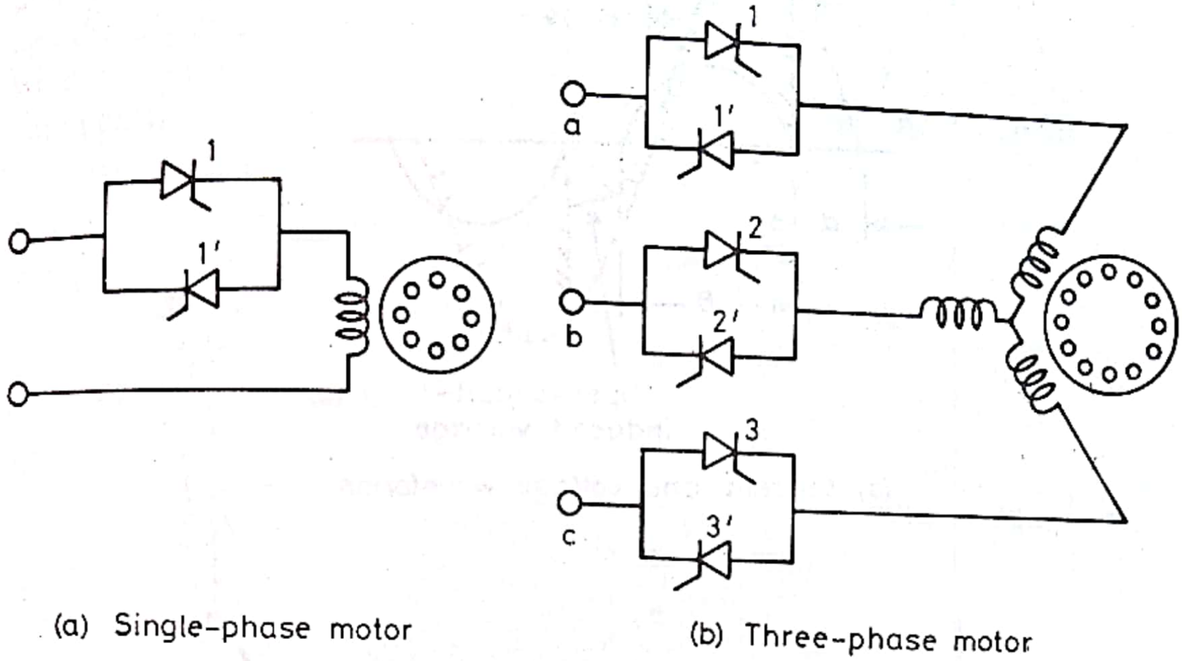


Fig. 6.11 Phase control of an AC motor.

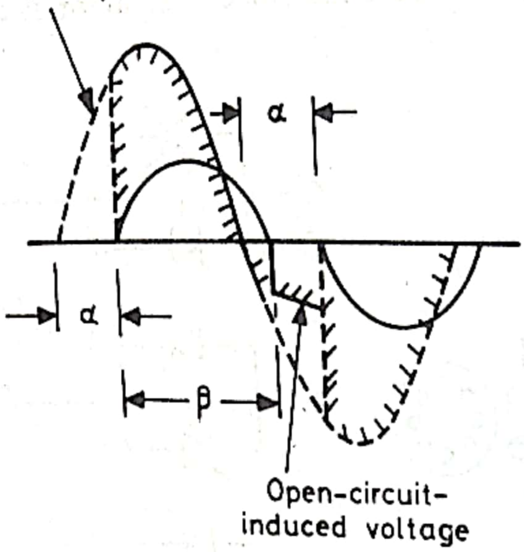
For single-phase circuits, the UJT relaxation oscillator shown in Fig. 6.9 can be used. The input current waveform is shown in Fig. 6.12a. The motor winding will experience open-circuits in every half-cycle if the angle of conduction β is less than π. During this period, the rotor currents will induce a voltage in the stator phase winding. The area shown by hatched lines in Fig. 6.12a is the motor phase voltage. As the firing angle increases, the RMS value of this voltage will decrease. The characteristics of an induction motor with variable applied voltage are shown in Fig. 6.12b. This method of control is very simple and economical. It provides a wide range of speed control if the load torque increases with speed as shown by curve 1 in Fig. 6.12b. The points indicated by circles on this curve show the several speeds that are possible by varying the voltage. If the load torque is constant, the speed variation is very much limited as shown by curve 2. Another drawback of this method of control is that the efficiency falls off with decrease in speed. For an induction motor, the power output is given by

$$\text{mechanical power output} = \text{power input to rotor} \times (1 - S),$$

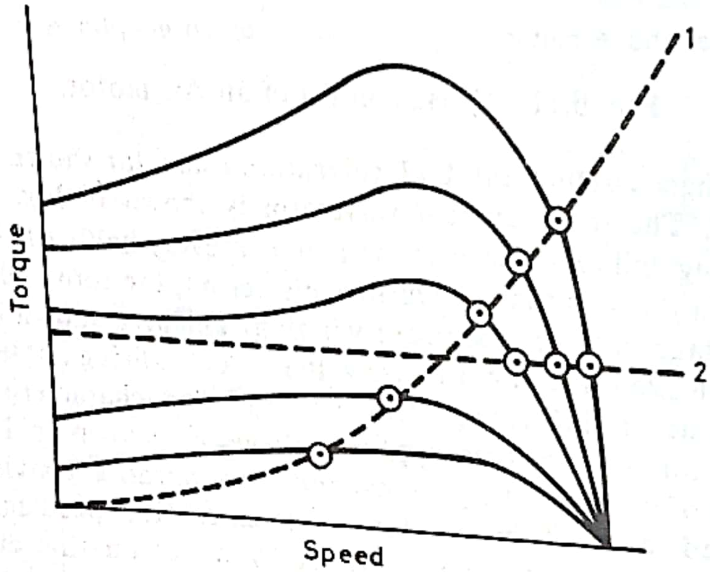
where S is the slip of the motor. Therefore,

$$\text{efficiency} = (1 - S). \tag{6.12}$$

If the load torque varies as a square of the speed, it can be shown that the motor takes maximum current at $S = 1/3$. If the speed reduction is obtained through phase control, then the peak current will increase by an additional 14% [see Paice, D. A., Induction motor speed control by stator voltage control, *IEEE Trans.* (PAS), 1968, p. 585].



(a) Current and voltage waveforms



(b) Speed-torque characteristics

Fig. 6.12 Characteristics of a phase-controlled AC motor.

6.6.1 PHASE CONTROL OF THREE-PHASE INDUCTION MOTORS

The schematic diagram of the phase-control circuit for controlling the speed of a three-phase induction motor is shown in Fig. 6.11b. The speed-torque characteristics and the overall performance are similar to those for a phase-controlled single-phase motor. The SCRs have to be triggered in a sequence to obtain balanced phase currents. The firing angle must be the same for all the phases. Figures 6.13a–c show the current waveforms for three firing angles. It will be observed that the motor exhibits three different modes of operation. When the conduction angle β is more than $2\pi/3$, the number of SCRs conducting at any time will be either 3 or 2. This is called the 3/2 mode. The phases get open-

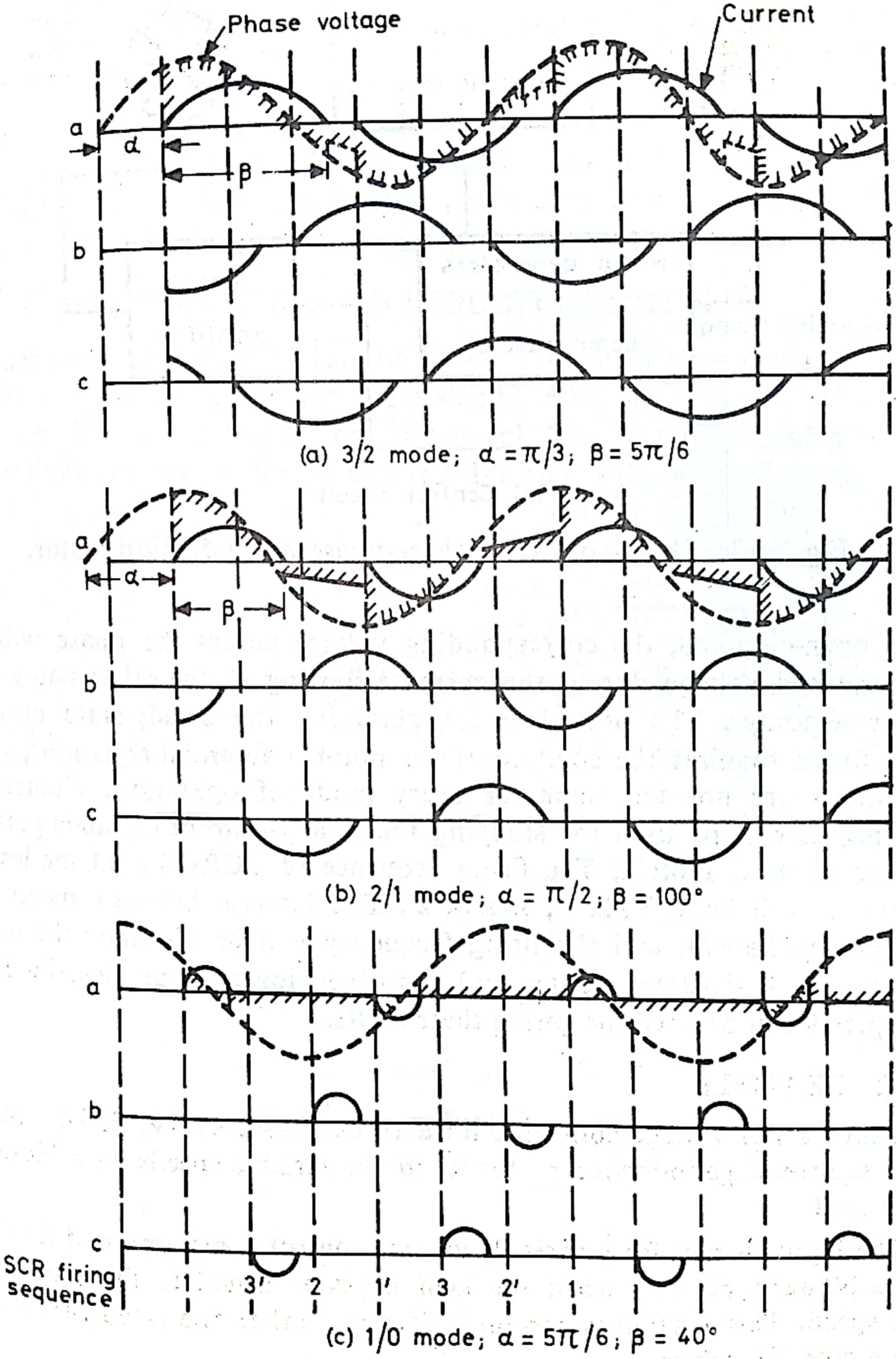


Fig. 6.13 **Phase-controlled three-phase AC induction motor** (*cont.*).

6.7 REGULATED DC POWER SUPPLIES

An important application of the controlled bridge circuits, discussed in Sections 6.1 and 6.2, is for regulated DC power supplies. A large series inductor and a shunt capacitor are used for reducing the ripple in the output. A feedback arrangement similar to that shown in Fig. 6.9 can be used for automatic adjustment of the firing angle of SCRs to control the output DC voltage. Figure 6.15 shows a commonly-used rectifier with an inductor input filter. A single-phase uncontrolled full-wave rectifier

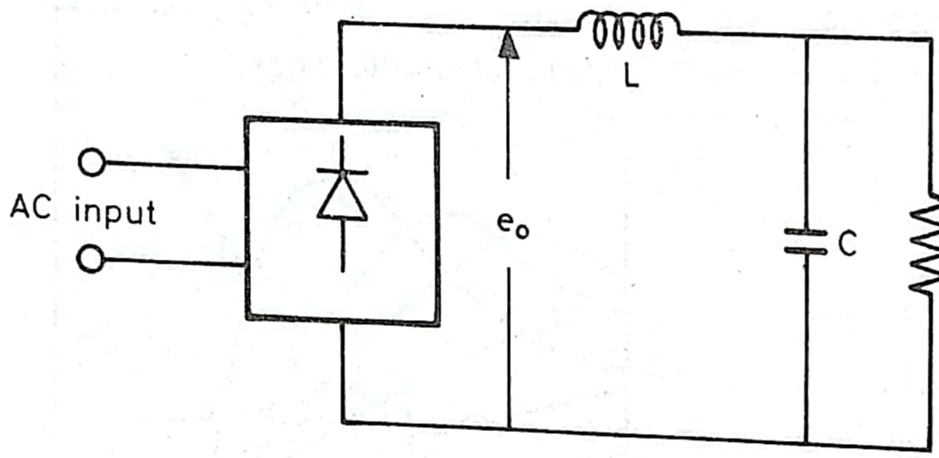


Fig. 6.15 Single-phase full-wave rectified power supply.

is considered for the filter design. If E_s is the effective value of the AC input, the Fourier series expansion for the rectified output voltage waveform is given by

$$e_o = \sqrt{2}E_s\left[\frac{2}{\pi} - \frac{4}{3\pi}\cos 2\omega t - \frac{4}{15\pi}\cos 4\omega t - \ldots\right]. \tag{6.13}$$

The ripple factor RF for the filtered output is defined as

$$\text{RF} = \frac{\text{RMS value of the harmonic voltages at the output}}{\text{DC voltage output}}.$$

For a single-phase full-wave bridge, we have

$$\text{RF} = \frac{1}{6(2)^{1/2}\omega^2 LC}, \tag{6.14}$$

where ω is the frequency of the input. Harmonics of an order higher than the second are neglected.

To maintain continuous conduction, the average direct current I_d must be at least equal to the peak amplitude of the second harmonic current through the inductor. If this condition is maintained by imposing a limit on the maximum value for load resistance R, then the DC output voltage will be approximately constant for all load currents. This criterion will be satisfied by

$$R_{max} = 3\omega L. \tag{6.15}$$

Equations (6.14) and (6.15) can be used for designing the required values of L and C for given RF and R_{max}. As the load resistance is increased beyond R_{max}, the load current will become discontinuous and the output voltage will rise above the average value $2(2)^{1/2}E_s/\pi$ and reach $(2)^{1/2}E_s$ when the load current is zero. A similar operation will take place when a controlled bridge is used.