# 6 Static drives and energy saving

## Contents

## 6.1 Speed control in squirrel cage motors

Speed control in slip-ring motors has been discussed in the previous chapter. Squirrel cage motors have limitations in their speed control in view of their fixed rotor parameters. Speed variation, in fixed steps, however, is possible in such motors if the stator is wound for multi-poles and such motors are known as pole changing motors. Up to four different speeds can be achieved in such motors economically, in combinations of 2/4, 4/6, 4/8, 6/8, 6/12, 2/4/6, 4/6/8, 2/4/6/12 and 4/6/8/12 poles etc. or any other similar combination. For limitation in the motor size and flux distribution, winding sets of more than two are not recommended. The two windings can be arranged for two, three or (maximum) four different speeds.

### 6.1.1 One winding

The single winding can be connected in delta/double star ($\Delta/YY$) to give two combinations of poles in the ratio of 2:1, i.e. 4/2, 8/4 or 12/6 poles etc. as shown in Table 6.1. Consequent poles like 4/6, 6/8, 8/12 motors can also be made with one winding.

### 6.1.2 Two windings

When more than two or non-multiple speeds are required (e.g. 4/6 or 6/8 etc.) then two windings can also be used. Each winding can further be connected in $\Delta/YY$ as noted above to give one additional speed for each winding and can thus be arranged for three or four different speeds as shown in Table 6.1.

## 6.2 Speed control through solid-state technology

In the following text we discuss how, with the application of varying supply parameters ($V$ and $f$), one can alter the characteristics of a fixed parameter induction motor in any desired way. We then deal with the application of solid-state technology to obtain the variations in the fixed supply parameters to achieve the required controls in an a.c. machine.

The static drives also provide a few more advantages such as

1 They transform an unbalanced supply system automatically to a balanced supply system through the switching logistics of the IGBTs* or the SCRs*. The feature is termed dynamic phase balancing.

*Note*
Wherever IGBTs or thyristors are mentioned in this book they would mean the latest versions in their respective families and are discussed in Sections 6.7.2 and 6.7.3.

2 The starting inrush current can be kept moderate for all types of drives, that can economize not only on ratings of the switchgears and cables but also on the size of the generator when a captive power is used to feed the load.

3. It is possible to operate a motor in both directions eliminating the use of an additional reversing starter.

The method for speed control as discussed earlier are only conventional and can only provide a discrete speed variation say, from 3000 r.p.m. to 1500 r.p.m. to 750 r.p.m. or vice versa. They cannot provide a smoother speed variation between any two speeds. The application of variable voltage is also not practical nor advisable, for it means a poor performance by the machine at lower voltages, whereas a higher voltage (more than 5% of the rated) is not permissible. Moreover, through this method the speed can be varied only within a very limited span due to very unstable conditions below the $T_{po}$ region, and a more than proportionate reduction in the h.p. developed. The torque also reduces in square proportion of the voltage. For such applications, therefore, which required smoother speed variation and over a wide range, one had no choice but to select d.c. drives. These drives were costly and needed higher maintenance because of commutators, slip-rings and brushes etc. that caused continuous arcing and required constant checks and maintenance. Also more down-time, which a process industry could least afford. Therefore combined systems were used but that required very elaborate arrangements, using two or more a.c. machines, rendering the whole system very cumbersome, vulnerable and yet more expensive. Since the speed was normally changed through the variation in frequency ($N \propto f$), these systems were basically frequency changers (converters) and were known as

- Cascade connections (concatenation)
  Use of two motors and the prime mover was a slip-ring motor.
- Frequency converters.
- Schrage type motor (commutator brush shifting arrangement).
- Leblanc or Scherbius Advancers.

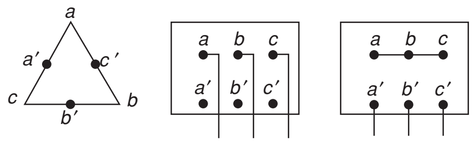
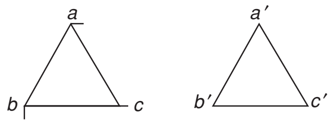
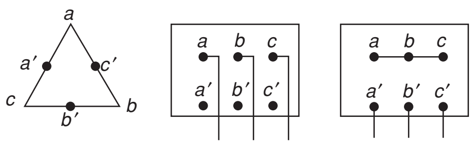
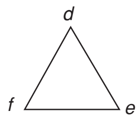
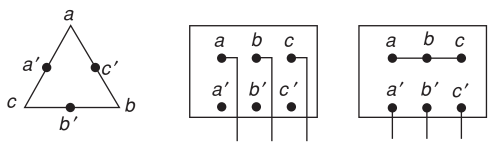
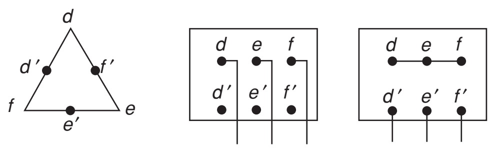
These systems were evolved to provide a variable – frequency supply source to feed directly the stator terminals of the a.c. motor or its rotor through the slip-rings. The motors had to be invariably a combination of two or more slip-ring motors to receive the rotor frequency voltage from the other machine or feedback the rotor frequency voltage to another machine. The easiest method was to have a variable-frequency supply source, which was not possible, unless the supply source itself was captive and earmarked for this drive alone or a combination of these drives on the same bus.

There was thus a practical limitation in employing an a.c. motor for all such applications that required frequent speed variation. Since these drives are no longer in practice, we have not considered it relevant to provide more details of these systems.

The above methods provide speed variation either in steps, as in squirrel cage motors or call for two machines or more, as in frequency converters, and cannot be used for a process line requiring frequent and precise speed controls. Until a few years ago there was no other option for all such applications except to use d.c. motors. D.C.

*IGBTs – Insulated gate bipolar transistors
SCRs – Silicon-controlled rectifiers (thyristors)

**Table 6.1** Connection diagrams for multispeed motors

| *No. of wdgs* | *Poles* | *Connection* | *Connection diagram* |
|---|---|---|---|
| One | 8/4 or 4/2 etc., also 6/8 or 4/6 etc. | Delta/double star |  Motor winding (i); 8-pole (ii); 4-pole (iii) |
| Two | 8/6 or 6/4 etc. | Star/star or delta/delta |  8-pole (i); 6-pole (ii) |
| Two | 8/6/4 or 6/4/2 etc. | One delta/double star and one star or delta |  Winding no. 1 8/4-pole (i); 8-pole (ii); 4-pole (iii)  Winding no. 2 6-pole (iv) |
| Two | 12/8/6/4 etc. | Two delta/double star |  Winding no. 1 12/6-pole (i); 12-pole (ii); 6-pole (iii)  Winding no. 2 8/4-pole (iv); 8-pole (v); 4-pole (vi) |

motors possess the remarkable ability of precise speed control through their separate armature and field controls. In d.c. motors the speed control below the base speed can be achieved through the armature control at constant torque and above the base speed through the field control at constant h.p. (Figure 6.7(b)). But a d.c. machine also has a few limitations:

1 Frequent maintenance due to continuously rubbing brushes mounted on a commutator.
2 Continuous arcing as a result of the above, give rise to fire hazards, particularly at installations that are contaminated with explosive gases, vapour or volatile liquids or are handling materials that are hazardous.
3 The maximum size of d.c. motor is limited due to limitation of maximum voltage across commutator segments for satisfactory commutation.

An induction motor, particularly a squirrel cage motor, is cheap, robust and is devoid of any such operating limitations and, has an obvious advantage over d.c. machines. It alone can provide an ultimate answer to such limitations. With the advent of static technology as discussed later, it has now become possible to make use of cage motors with the same ease and accuracy of speed control and even better than d.c. machines. Static drives respond extremely fast as they can be microprocessor based. They can compute process data and provide system corrections almost instantly (called 'realtime processing') as fast as within 1–2 ms and even less. In Table 6.5 we show a broad comparison between a d.c. machine and a static drive using cage motors. It gives an idea of applying static technology to all process requirements with more ease and even better accuracy. With the advent of this technology, the demand for d.c. machines is now in decline as noted in Section 6.19.

### 6.2.1 Theory of application

The application of solid-state technology for the speed control of a.c. motors is based on the fact that the characteristics and performance of an induction motor can now be varied, which until a few years ago were considered fixed and uncontrollable. With the advent of solid-state technology, which was introduced around 1970 for industrial applications, the motor's parameters and therefore its performance can now be varied by varying the supply parameters of the system. For example the voltage and frequency in cage motors, and rotor resistance or rotor current in slip-ring motors, as discussed in Chapter 1. This technology can also provide a varying resistance in the rotor circuit of a slip-ring motor by varying the rotor current as discussed in Section 6.16.3 without the loss of power in the external resistance. It is thus also suitable to provide speed control in a slip-ring motor. Speed control of slip-ring motors with the use of solid-state technology is popularly known as slip recovery systems, as the slip power can also be fed back to the source of supply through a solid-state feedback converter bridge, discussed later.

### 6.2.2 Effects of variable supply parameters on the performance of an induction motor

Here we analyse the effect of variation in the incoming supply parameters (voltage and frequency) on the characteristics and performance of an induction motor (such as its flux density, speed, torque, h.p., etc). We also assess the effect of variation of one parameter on the other, and then choose the most appropriate solid-state scheme to achieve a required performance. We generally discuss the following schemes:

1 *V/f* or v.v.v.f. (variable voltage variable frequency) control (speed control at constant torque)
2 Phasor (vector) control
   – Single-phasor (vector) control
   – Field-oriented control (FOC), commonly known as double-phasor or phasor (vector) control and
   – Direct torque control (DTC).

## 6.3 *V/f* or v.v.v.f. control (speed control at constant torque)

This is also known as variable frequency control. Consider the following equations from Chapter 1:

$$T \propto \frac{S \cdot {}_{ss}e_2^2 \cdot R_2}{R_2^2 + S^2 \cdot {}_{ss}X_2^2} \qquad (1.3)$$

and $T \propto \phi_m \cdot I_{rr}$ (1.1)

and $e_2 = 4.44\ Kw \cdot \phi_m \cdot Z_r \cdot f_r$ (1.5)

$\therefore$ for the same supply voltage $V_1$

$e_2 \propto \phi_m \cdot Z_r \cdot f_r$

i.e. $\phi_m \propto \dfrac{e_2}{f_r}$

and $T \propto \phi_m \propto \dfrac{e_2}{f_r}$

i.e. for the same design parameters ($\phi_m$ remaining the same) and ratio $e_2/f_r$, the torque of the motor, $T$, will remain constant. Since both $e_2$ and $f_r$ are functions of the supply system, a variation in $V_\ell$ and $f$ can alter the performance and the speed–torque characteristics of a motor as required, at constant torque. By varying the frequency smoothly from a higher value to a lower one or vice versa (within zero to rated), an almost straight line torque can be achieved (Figure 6.3). This type of a control is termed variable voltage, variable frequency (v.v.v.f or *V/f*) control. At speeds lower than rated, the natural cooling may be affected, more so at very low speeds, and may require an appropriate derating of the machine or provision of an external or forced cooling. The practice of a few manufacturers, up to medium sized motors, is to provide a cooling fan with separate power connections so that the cooling is not affected at lower speeds.

*Note*
The speed of the motor can be varied by varying the frequency alone but this does not provide satisfactory performance. A variation in frequency causes an inverse variation in the flux, $\phi_m$ for the same system voltage. The strength of magnetic field, $\phi_m$, develops, the torque and moves the rotor, but at lower speeds, $f$ would be reduced, which would raise $\phi_m$ and lead the magnetic circuit to saturation. For higher speeds, $f$ would be raised, but that would reduce $\phi_m$, which would adversely diminish the torque. Hence frequency variation alone is not a recommended practice for speed control. The recommended practice is to keep *V/f* as constant, to maintain the motor's vital operating parameters, i.e. its torque and $\phi_m$, within acceptable limits.

The above is valid for speed variation from zero to the rated speed. For speed variations beyond the rated speed, the theory of *V/f* will not work. Because to maintain the same ratio of *V/f* would mean a rise in the applied voltage which is not permissible beyond the rated voltage, and which has already been attained by reaching the rated speed. The speed beyond the rated is therefore obtained by raising the supply frequency alone (in other words, by weakening the field, $\phi_m$) and maintaining the voltage as constant at its rated value. We can thus achieve a speed variation beyond the rated speed by sacrificing its torque and maintaining the product $T \cdot f$ as constant. Since $P \propto T \cdot N_r$ and $N_r \propto f_r$, therefore speed variation can now be achieved at constant power output (see Figure 6.4). We have combined Figures 6.3 and 6.4 to produce Figure 6.5 for more clarity, illustrating the speed variation at constant torque below the base speed and at constant h.p. above the base speed.

*V/f* is the most commonly used method to control the speed of a squirrel cage motor. The fixed frequency a.c. supply, say, at 415 V, 50 Hz from the mains, is first rectified to a constant or variable d.c. voltage, depending upon the static devices being used in the inverter circuit and their configuration. This voltage is then inverted to obtain the required variable-voltage and variable-frequency a.c. supply. The basic scheme of a *V/f* control is illustrated in Figure 6.6. The approximate output voltage, current and torque waveforms are shown in Figure 6.7(a). The torque curve is now almost a straight line, with only moderate pulsations, except for some limitations, as discussed later, enabling the drive to start smoothly. When, however, a higher starting torque is required to start the motor quickly, it is also possible to boost the starting torque to a desirable level (up to the designed $T_{st}$ of the motor) by raising the voltage to $V_r$, through the pulse width modulation (PWM), discussed in Section 6.9.2. The starting current can also be reduced to only 100–150% of the rated current or as desired, to the extent possible, by suitably varying the *V/f*. The control can thus also provide a soft-start switching. It may, however, be noted that there is no control over the starting current, which is a function of the applied voltage, and the minimum voltage during start-up will depend upon the motor, the load characteristics and the thermal withstand time of the motor (Section 2.8). With the use of static control drives, it is however, possible to minimize the starting inrush current to a reasonable level but with a loss of starting torque. To do so, it is advisable to match the load and the motor starting characteristics and determine the minimum starting torque required to pick up the load. For this starting torque is then adjusted the starting voltage. The magnitude of the starting current will then depend upon this voltage.

The static drives too are defined by their over-current capacity and its duration. IEC 60146-1-1 has defined it as 150–300%, depending upon the type of application, for a maximum duration of one minute. For instance, for centrifugal drives, which are variable torque drives (Figures 2.10 and 2.11) such as pumps, fans and compressors etc. an over-current capacity of the order of 110% for one minute would be adequate. For reciprocating drives, which are constant torque drives (Figures 2.12 and 2.13), such as ball mills and conveyors, a higher over-current capacity say, 150–300% for one minute, would be essential. For higher starting currents or longer durations of start, the drive may have to be derated, which the manufacturer of the drive alone will be able to suggest. Drive ratings are basically selected on the basis of motor rating and the starting currents and are rated for a starting current of 150% for one minute. If more than 150% is required, then the rating of the drive is worked out on the basis of the starting current requirement and its duration. *V/f* is a concept to vary the speed, maintaining a constant torque. Through a static drive, however, it is possible to vary one parameter more than the other, to obtain any speed–torque characteristic to meet a particular duty cycle. By varying the rectifier and inverter parameters, a whole range of *V/f* control is possible. High-inertia loads, calling for a slip-ring motor for a safe and smooth start, can now make use of a standard squirrel cage motor. Open- or closed-loop control systems can be employed to closely monitor and control the output voltage and frequency so that the ratio of *V/f* is always maintained constant.

## *Limitations of V/f control*

Figure 6.1 illustrates theoretical speed–torque curves. In fact at very low speeds, say, at around 5% of $N_r$ (at 5% *f*) or less, the motor may not be able to develop its theoretical torque due to a very low stator voltage, on the one hand, and relatively higher proportion of losses and a lower efficiency of the machine, on the other. Figure 6.2 illustrates a realistic torque characteristic at different supply frequencies. They display a sharply drooping and rather unstable performance at very low speeds. For more accurate speed controls at such low speeds, one may have to use phasor-controlled drives, discussed later, or cyclo-converters, which are relatively very costly and are used, for very large drives. Phasor-controlled drives are available at reasonable cost and can provide extremely accurate speed controls even at very low speeds, 5% of $N_r$ and less. Typical applications calling for such a high torque at such low speeds could be a steel plant process line or material handling (e.g. holding a load stationary at a particular height by the crane, while shifting material from one location to another).

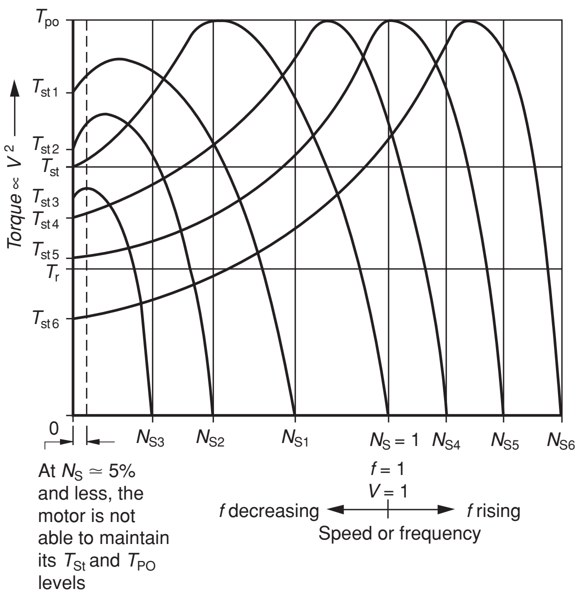



**Figure 6.1** Approximate theoretical representation of speed versus torque for *V/f* control of a motor at different values of *f*

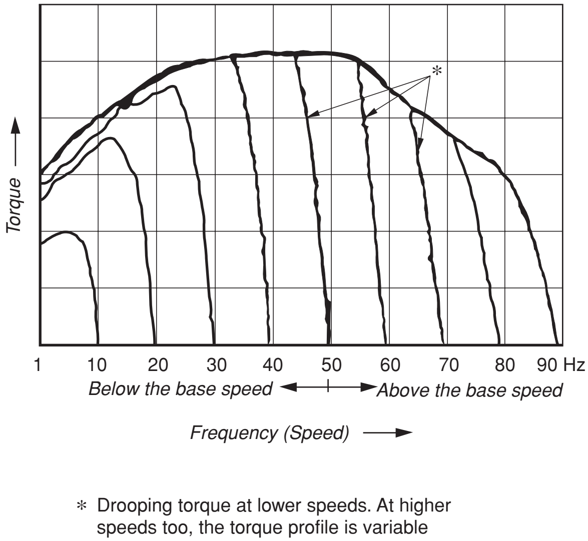



**Figure 6.2** Actual speed–torque characteristics by a conventional frequency control (*V/f* control)

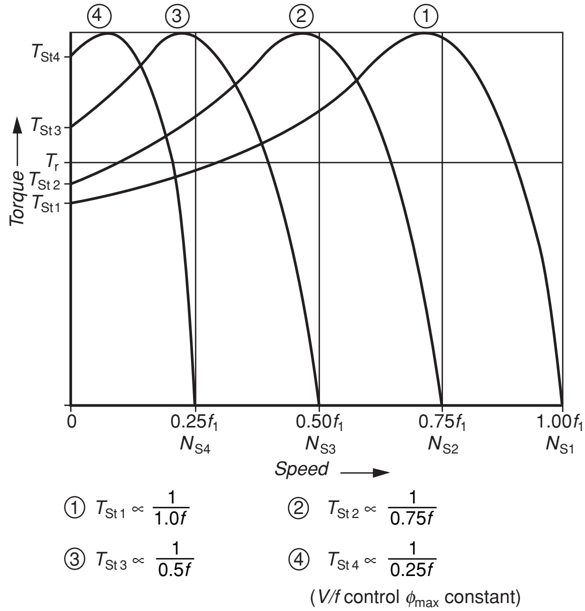



**Figure 6.3** Speed variation at constant torque

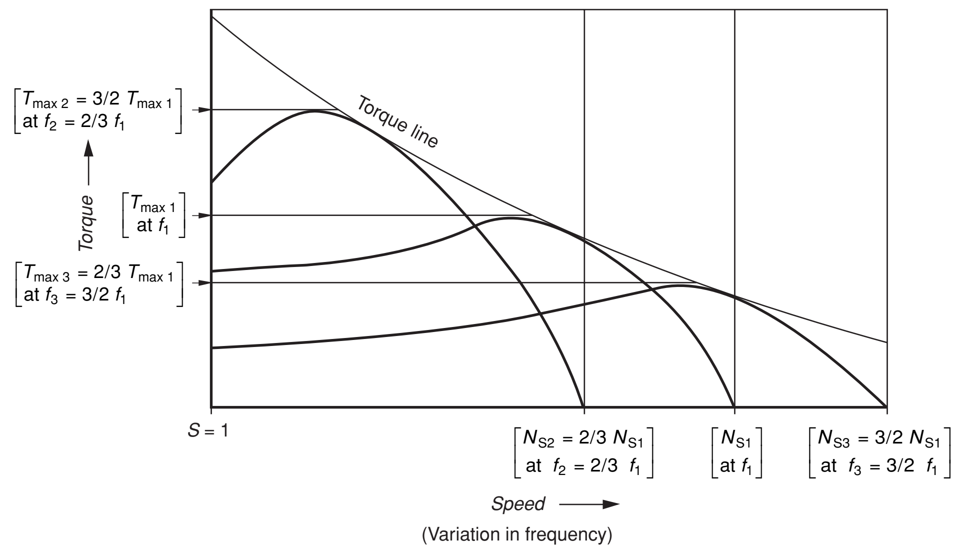



**Figure 6.4** Speed variation at constant HP

### *Application*

In a *V/f* control generally, only the frequency is varied to obtain the required speed control. Based on this frequency, the switching logistics of the inverter control circuit control the inverter's output voltage using the PWM technique to maintain the same ratio of *V/f*. A *V/f* control is, however, not suitable at lower speeds. Their application is limited to fan, pump and compressor-type loads only, where speed regulation need not be accurate, and their low-speed performance or transient response is not critical and they are also not required to operate at very low speeds. *V/f* controls are primarily used for soft starts and to conserve energy during a load variation (see Example 6.1 for more clarity). Rating, however, is no bar.

## 6.4 Phasor (vector) control

A simple *V/f* control, as discussed above, will have the following limitations:

- Control at very low speeds is not possible.
- Speed control may not be very accurate.
- Response time may not be commensurate with the system's fast-changing needs.

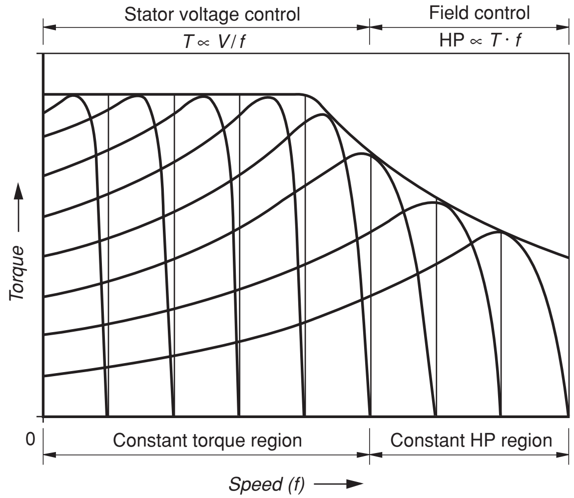



**Figure 6.5** Speed control in an a.c. motor

All these parameters are extremely essential for a process line. With the R&D in the field these limitations have been overcome with the use of phasor controls. To implement these controls different manufacturers have adopted to different control and feedback systems to monitor and control the torque and field components. They have also given these controls different trade names. While the basic technological concept may remain the same, process implementation may vary from one manufacturer to another. Below we attempt to identify the more common phasor controls introduced by a few leading manufacturers.

## 6.4.1 Single phasor (vector) control

Let us consider the simple equivalent motor circuit diagram as shown earlier in Figure 1.15. The no-load component of the current, $I_{n\ell}$, that feeds the no-load losses of the machine contains a magnetizing component, $I_m \cdot I_m$ produces the required magnetic field, in the stator and the rotor circuits, and develops the rotor torque so that

$$T \propto \phi_m \cdot I_{rr} \tag{1.1}$$

The magnetizing current, $I_m$, is a part of the motor stator current $I_\ell$ (Figure 1.15). The rotor current is also a reflection of the active component of this stator current, as can be seen in the same figure, so that

$$\bar{I}_\ell = \bar{I}_{n\ell} + \bar{I}_a$$

$$= \bar{I}'_m + \bar{I}_m + \bar{I}_a \tag{6.1}$$

All these are phasor quantities. $I_a$ is the active component responsible for developing the rotor torque and $I_m$ the magnetic field. Varying $I_a$ would mean a corresponding variation in the torque developed.

### *Variation of speed below the base (rated) speed*

The machine now operates in a constant torque region (Figure 6.7(b)). The frequency is reduced as is the voltage to maintain the same ratio of *V/f*. At lower voltages, $I_\ell$ and therefore $I_a$ will diminish, while $\phi_m$ and $I_m$ will rise, so that $\phi_m \cdot I_{rr}$ is a constant.

Equation (6.1) can be rewritten, for better analysis with little error as

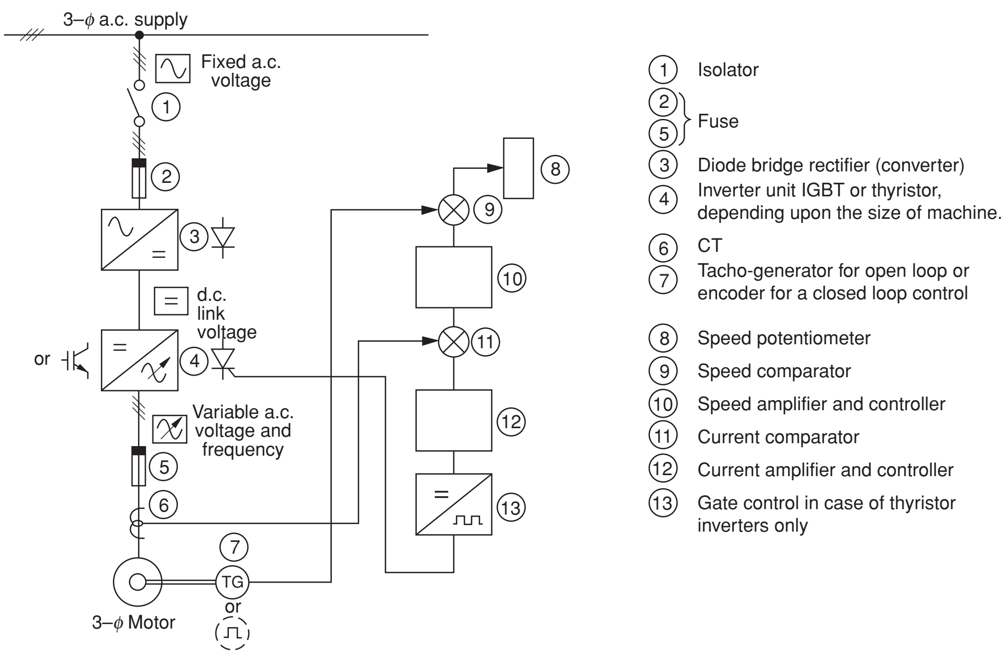



**Figure 6.6** Typical block diagram of a *V/f* control scheme with open- or closed-loop control scheme

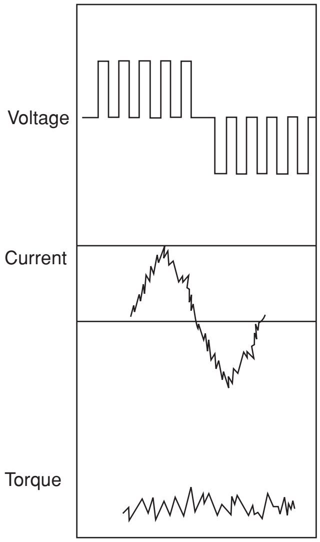



**Figure 6.7(a)** Approximate characteristics of vital parameters after pulse width modulation

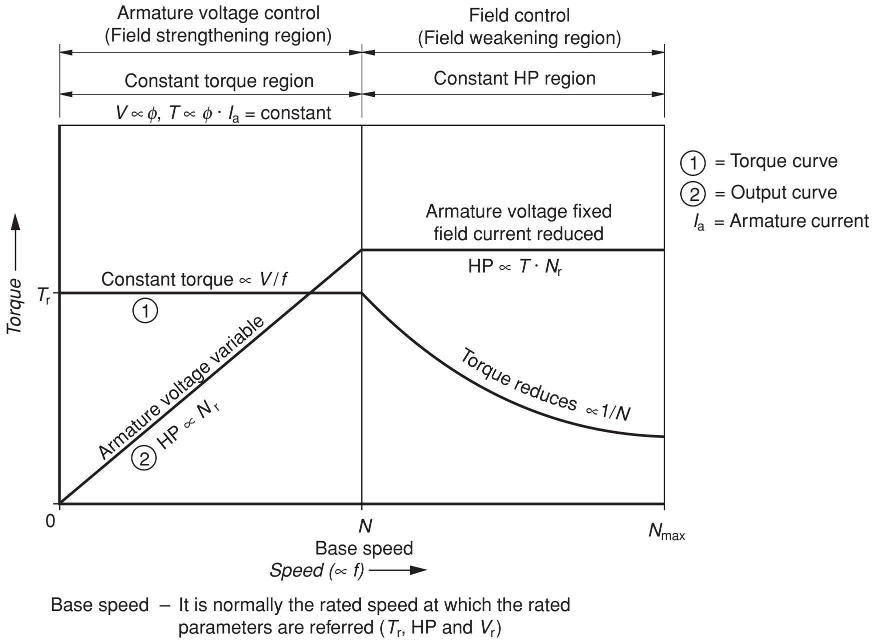



**Figure 6.7(b)** Variation of torque with speed in a d.c. machine (same for an a.c. machine)

$$I_\ell \simeq \bar{I}_a + \bar{I}_m$$

where,

$I_a \propto T \propto V_1$ and

$I_m \propto f$

Therefore a normal *V/f* can also be transformed into a phasor control, $\bar{I}_a$ and $\bar{I}_m$ being torque- and flux-producing components respectively. These components are represented only theoretically. In fact they are not separate and hence the difficulty in controlling each of them more precisely.

### *Variation of speed beyond the base (rated) speed*

See Figure 6.7(b). The machine now operates in a constant h.p. region. The frequency is raised but the voltage is kept constant at its rated value (as it should not be raised beyond rated). The flux will diminish while $I_\ell$ and also $I_{rr}$ will remain almost the same. The torque therefore reduces so that the h.p. developed remains almost a constant (h.p. $\propto T \cdot N$). This is also known as the field weakening region.

In the above schemes the two quantities ($I_a$ and $I_m$) are not separated. Initially it was not easy to separate them and the whole phasor $I_\ell$ was varied to achieve a speed variation. Yet close speed control was possible but the motor's basic parameters were essential to achieve more accurate speed control. Since it may not be practical to obtain all the parameters of each motor promptly, the drive software is designed so that the name plate particulars of a motor ($V$, $I_r$, $N_r$, and $kW$) alone are enough to determine the machine's required parameters through the motor's mathematical model. The motor is calibration* run on no-load at different voltages and speeds and the drive is able to establish near-accurate vital parameters of the machine in terms of the equivalent circuit diagram shown in Figure 1.15 earlier. With these parameters known, it is now possible to achieve the required speed controls noted above. Since these parameters are fixed for a motor, the motor has to be selected according to the load duty and may require a pre-matching with the load.

### *Application*

Such a control is good for machines that are required to operate at low speeds with a high accuracy. Now the phasor $I_\ell$, in terms of $I_m$, is varied according to the speed required. Figure 6.2 now changes to Figure 6.8, which is a marked improvement on the earlier characteristics. The torque variation with speed is now almost constant, except at very low speeds. The reason for poor torque at low speeds is the method of speed variation which is still based on *V/f*. Now a motor's mathematical model is used to vary the speed of the machine by sensing the component, $I_m$, of the machine. Any variation in the actual $I_m$ than the desired pre-set value in the inverter switching logistics

*1 Calibration run or auto-tuning is a feature of the motor's mathematical model that can establish the motor parameters with its test run.
2 If the actual motor data are available from the motor manufacturer, a calibration run will not be necessary. The motor's mathematical model can now be fed with the explicit data to achieve more precise speed control.
3 All the motor models are implemented by microcomputer software.

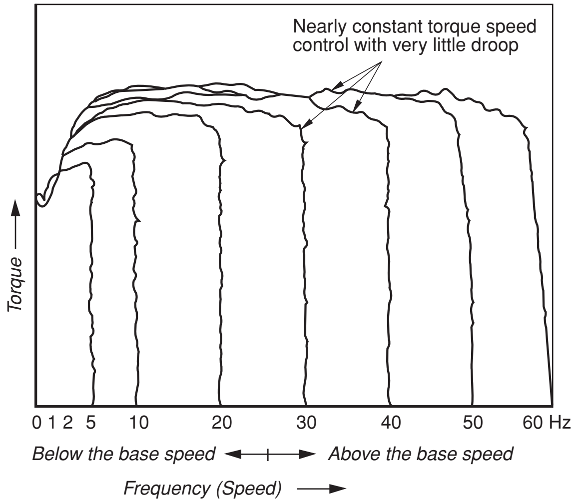



**Figure 6.8** Speed–torque characteristics by flux ($I_m$) control (single phasor control)

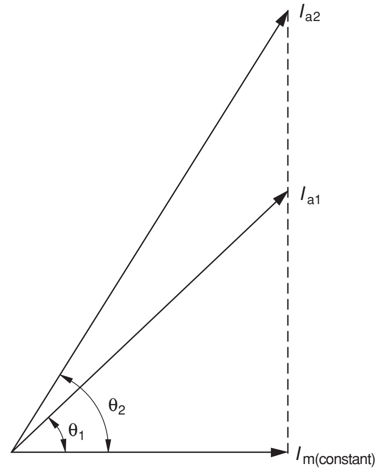



**Figure 6.9** Phasor representation of field current ($I_m$) and stator active current ($I_a$)

is made up by the PWM technique. The field-oriented block diagram illustrated in Figure 6.12 can be suitably simplified for $I_m$ control. Tachogenerator or pulse encoder feedback devices can be employed to achieve higher accuracy in speed control.

With years of research and development in the field of static drives, it is now possible to identify and separate these two parameters ($I_a$ and $I_m$) and vary them individually, as in a d.c. machine, to achieve extremely accurate speed control, even slightly better than in d.c. machines. In d.c. machines the armature current and the field strength are also varied independently. A.C. machines can now be used to provide very precise speed control, as accurate as ± 0.001% of the set speed, with closed-loop feedback controls. This technique of speed control is termed field-oriented control (FOC) and is discussed below.

### 6.4.2 Field-oriented control (FOC)

This is commonly known as double phasor or phasor (vector) control. If we analyse Equation (1.1) in Chapter 1 we will observe that $\phi$ is a function of stator magnetizing current, $I_m$, and $I_{rr}$ is a transformation of the stator active current, $I_a$.

Hence, Equation (1.1) can be rewritten as

$$T \propto \bar{I}_m \cdot \bar{I}_a$$

Both of them are phasor quantities, and are shown in Figure 6.9. In absolute terms, they can be represented by

$$T = k \cdot I_m \cdot I_a \sin \theta \qquad (6.2)$$

where $\theta$ represents the electrical position of the rotor field in space with respect to the stator. In other words, it is the phasor displacement or slip angle between $\bar{I}_m$ and $\bar{I}_a$ and will continue to vary with variation in load (torque). For instance, referring to Figure 6.9, the smaller the load $I_{a1}$, the smaller will be $\sin \theta_1$ and the larger the load $I_{a2}$, the larger will be $\sin \theta_2$. Thus to achieve a required level of speed control the stator current, $I_a$, field current, $I_m$, and phasor angle, $\theta$, can be suitably varied. Since it is the phasor of the rotor flux (rotating field), i.e. the magnitude and its angular position with respect to the active current of the stator, which is being varied, to achieve the required speed control, this phasor control is called field oriented control (FOC). The theory of field orientation was first introduced by F. Blascke in 1972 (see Blascke (1972) and *EPE Journal* (1991)). Having been able to identify the rotor field phasor it is now possible to vary this and obtain a speed control in a squirrel cage machine similar to that in a d.c. machine.

For field-oriented controls, a mathematical model of the machine is developed in terms of rotating field to represent its operating parameters such as $N_r$, $I_a$, $I_m$ and $\theta$ and all parameters that can influence the performance of the machine. The actual operating quantities are then computed in terms of rotating field and corrected to the required level through open- or closed-loop control schemes to achieve very precise speed control. To make the model similar to that of a d.c. machine, Equation (6.2) is further resolved into two components, one direct axis and the other quadrature axis, as discussed later. Now it is possible to monitor and vary these components individually, as with a d.c. machine. With this phasor control we can now achieve a high dynamic performance and accuracy of speed control in an a.c. machine, similar to a separately excited d.c. machine. A d.c. machine provides extremely accurate speed control due to the independent controls of its field and armature currents.

Different manufacturers have adopted different methods with minor changes to achieve almost the same objective. For example, field-oriented control was first introduced by Allen Bradley in the USA in 1981 and a similar technique was introduced at the same time by ABB of Finland. ABB claim their technique to be still faster in responding, as it eliminates the modulation section of the drive (which we will discuss later, when discussing drives), and the torque could be controlled directly. They call it direct torque control (DTC).

The phasor $\bar{I}_a$ and $\bar{I}_m$ are separated and then controlled separately as discussed later. For more precise speed control a pulse encoder feedback device can also be employed. The characteristics now improve to Figure 6.10. The torque can now be maintained constant at any speed, even at zero speed.

With different approaches to monitor and control the basic parameters of the motor, i.e. $I_a$, $I_m$ and $\sin\theta$, many more alternatives are possible to achieve the required speed variation in an a.c. machine. Control of these parameters by the use of an encoder can provide an accuracy in speed control as good as a d.c. machine and even better.

### *To implement the FOC*

The field phasor is a continuously rotating phasor in the space, whose angular position keeps changing with the position of the rotor with respect to the stationary stator. Let the rotor field displacement under the stationary condition with respect to the stator be denoted by angle $\beta$ as shown in Figure 6.11. This displacement will continue to change and will rotate the rotor (field frame). All the phasor quantities of the stator are now expressed in terms of the field frame. Figure 6.11 shows these two equivalent stator side phasors transformed to the rotor frame.

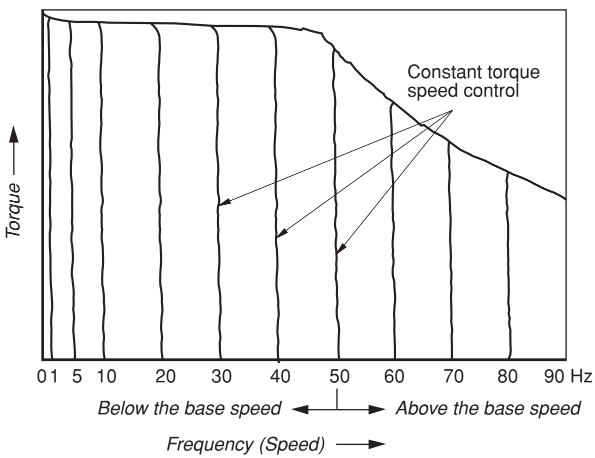



**Figure 6.10** Speed–torque characteristics by field-oriented control (FOC) (flux and torque control) (Source: Allen Bradley)

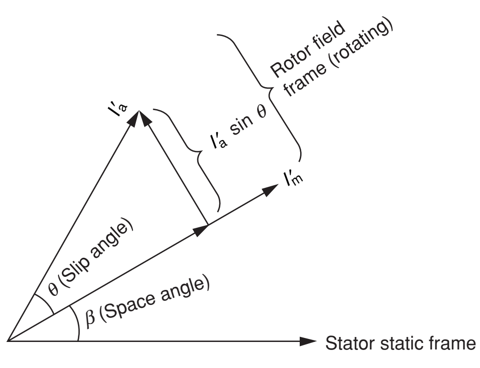



**Figure 6.11** Rotor field reference frame

$I'_a$ = corresponding stator phasor for active current, referred to the rotor side

$I'_m$ = corresponding stator phasor for the magnetizing current, referred to the rotor side

$\theta$ = phase displacement between the stator active and magnetizing current components

$\beta$ = angular displacement of the rotor field with respect to the stationary stator at a particular instant. It will continue to vary with the movement of the rotor.

The phasor diagram would suggest that:

1 $I'_a \sin\theta$ can be regarded as the quadrature component. It is the torque component on which will depend the torque developed by the rotor. It is this component that will be varied for speed variations below the base speed, maintaining the field current constant according to the rated condition. It is similar to the armature current control in a d.c. machine.

2 $I'_m$ can be regarded as the direct axis field component responsible for the field flux. This can be weakened (reduced) for speed variations above the base speed, which is the constant-output, constant-voltage region. Now the torque component will diminish. It is similar to the separately excited field control in a d.c. machine. Both these phasors can be regulated separately like a d.c. machine to achieve any speed variation with high precision and accuracy and provide a high dynamic performance. Leading manufacturers have developed mathematical models with microprocessors to determine the modulus and the space angle $\beta$ of the rotor flux space phasor through $I_a$ and speed. The space angle of the rotor flux space phasor is then obtained as a sum of slip angle $\theta$ and the field angle $\beta$. The slip angle $\theta$ can be calculated from the reference values $I_a$ and $I_m$ (an indirect method, as no sensors are used). With these parameters known, it is now possible to identify the position of the rotor flux phasor and then orientate the stator current phasor to determine the relative displacement between the two in the space, to achieve a phasor diagram as shown in Figure 6.11. The phasor diagram provides crucial parameters and must be established accurately to obtain accurate results from the control of $I'_a$ and $I'_m$. The relationship between these two phasors, $I'_a$ and $I'_m$, is then monitored closely and controlled by adjusting the supply parameters to the stator of the machine. The supply parameters can be controlled through a VSI (voltage source inverter) or CSI (current source inverter), which are discussed later, depending upon the practice of the manufacturer.

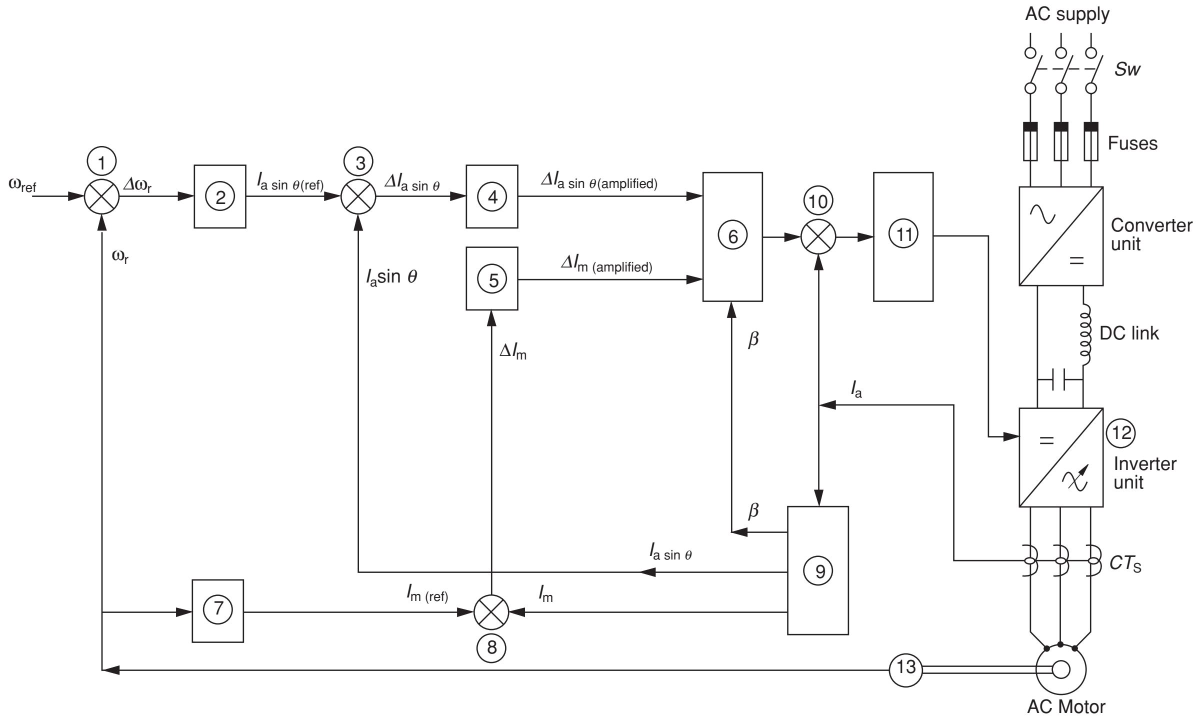
The microprocessor plays the role of an electronic controller that transforms electrical quantities such as *V*, *I* and *N* etc. into space flux phasors, to be compared with the pre-set data. It then creates back *V*, *I* and *N* etc., which are the controlling variables, in terms of correction required and feeds these to the machine through a VSI or CSI to achieve the required controls.

Mathematical modelling of the machine is a complex subject and is not discussed here. For this, research and development works carried out by engineers and the textbooks available on the subject may be consulted. A few references are provided in the Further Reading at the end of this chapter. In the above analysis we have considered the rotor flux as the reference frame. In fact any of the following may be fixed as the reference frame and accordingly the motor's mathematical model can be developed:

Rotor flux-oriented control – when the rotor is considered as the reference frame.
Stator flux-oriented control – when the stator is considered as the reference frame and
Magnetic field-oriented control – when the field is considered as the reference frame.

The rotor flux-oriented control is more popular among different manufacturers to achieve high precision of speed control in an induction machine. With this technology (any of the three methods noted above), it is now possible to obtain a high performance of the machine, i.e. torque up to 100% of $T_r$ at speeds down to zero.

Since the motor's fixed parameters can now be varied to suit a particular load requirement, there is no need to pre-match a motor with the load as was necessary in a single phasor control. Now any motor can be set to achieve the required characteristics to match with the load and its process needs. Full-rated torque ($T_r$) at zero speed (during start) should be able to pick-up most of the loads smoothly and softly. Where, however, a higher $T_{st}$ than $T_r$ is necessary, a voltage boost can also be provided during the start-up to meet this requirement. (See also Section 6.16.1 on soft starting.) The application of phasor (vector) control in the speed control of an a.c. motor is shown as a block diagram in Figure 6.12.




(1) Speed comparator: to determine the speed error ($\omega_{ref} - \omega_r$) where,
$\omega_{ref}$ = required speed reference and
$\omega_r$ = actual speed.

(2) Speed control amplifier to feed the reference torque data $I_a$ sin $\theta_{(ref.)}$ to the torque error detector. This is the quadrature component.

(3) Current comparator: It is the torque error detector, to detect the torque component error $\Delta I_a$ sin $\theta$ ($I_a$sin $\theta_{(ref)}$ – $I_a$ sin$\theta$).

(4) Torque error amplifier to feed the torque error to carry out the desired correction through block 6.

(5) Field error $\Delta I_m$ amplifier.

(6) Phasor rotator to transform the field frame coordinates to the stator frame coordinates.

(7) Field weakening unit to command the field strength $I_{m(ref.)}$ for speed regulation above the base speed.

(8) Field current comparator ($\Delta I_m$) to determine ($I_{m(ref)}$ – $I_m$).

(9) Motor flux mathematical model to determine prevailing $I_a$ sin $\theta$ and $I_m$ and also the space field displacement angle with respect to the stator '$\beta$'.

(10) Current comparator to compute the error between the actual line quantities and the desired quantities from block 6 and give command to the control unit 11.

(11) Switching block controls the switching of the inverter unit to regulate its output to the preset reference line quantities.

(12) Inverter unit.

(13) Pulse encoder – To feed back actual speed of the motor and the angular position of the rotor with respect to the stator at a particular instant.

**Figure 6.12** Block diagram for a flux-oriented phasor control

### *Application*

FOC drives are capable of providing precise speed control and are used for applications calling for high performance and precision (e.g. machine tools, high-speed elevators, mine winders, rolling mills, etc.). These drives are capable of regulating a number of variables at the same instant such as speed, position, acceleration and torque.

## 6.4.3 Direct torque control (DTC)

This is an alternative to FOC and can provide a very fast response. The choice of a static drive, whether through a simple *V/f* control, field-oriented phasor control or direct torque control with open or closed-loop control and feedback schemes, would depend upon the size of the machine, the range of speed control (whether required to operate at very low speeds, 5% $N_r$ and below), the accuracy of speed control and the speed of correction (response time). The manufacturers of such drives will be the best guide for the most appropriate and economical drive for a particular application or process line.

This technology was introduced by M/s ABB of Finland to achieve an extremely fast and highly accurate speed control in an a.c. machine. This is also based on phasor control, but the field orientation is now obtained without using a modulation (PWM) control circuit. The manufacturer makes use of only motor theory, through a highly accurate mathematical motor model, to calculate the motor torque directly. There is no need to measure and feed back the rotor speed and its angular position through an encoder to the motor model. The controllable variables now are only $\phi$ and $T$. It is therefore called a direct torque control (DTC) technique. There is no need to control the primary control variables $V$ or $I$ and $N$ now as in a flux phasor control. This scheme eliminates the signal processing time in the absence of a PWM and also an encoder circuit, both of which introduce an element of delay. As these drives control $\phi$ and $T$ directly, they respond extremely quickly and it is possible to achieve a response time as low as 1–2 ms. They are thus a corollary to the sensor-less flux control drives. For reference, a rough comparison between the various types of drives is given in Table 6.2.

### *Basic scheme of a DTC drive*

A simple block diagram as shown in Figure 6.13 illustrates the operation of a DTC drive. It contains two basic sections, one a torque control loop and the other a speed control loop. The main functions of these two control circuits are as follows:

### 1 Torque control loop

### Section 1

This measures

- the current in any two phases of the motor
- the d.c. bus voltage, which is a measure of the motor voltage
- the switching position of the inverter unit.

**Table 6.2** Comparison of conventional *V/f* control with different modes of phasor control drives

| Serial no. | Performance parameters | *Variable-frequency drives* | | *Phasor (vector) control drives* | | | | |
|---|---|---|---|---|---|---|---|---|
| | | *V/f control* | | *Flux ($I_m$) control (single-phasor control)* | *Double-phasor (vector) or field-oriented control (FOC)* | | *Direct torque control (DTC)* | |
| | | *Without encoder*[a] | *With encoder*[b] | *a, c* | *Without encoder*[a] | *With encoder*[b] | *Without encoder*[a] | *With encoder*[b] |
| 1 | Response time in adjusting the speed (regulation) | 200–400 ms | 100–200 ms | 125–250 ms | 10–20 ms | 5–10 ms | 1–2 ms | [d] |
| 2 | Accuracy of speed control (regulation) | ± 1.0% | ± 0.1% | ± 0.5% | ± 0.5% | ± 0.001% | ± 0.1% | ± 0.001% |
| 3 | Speed range | 40 : 1 | 40 : 1 | 120 : 1 | 120 : 1 | 1000 : 1 | 100 : 1 | 100 : 1 |
| 4 | Ability to maintain 100% rated torque at zero speed | Not possible | Not possible | Not possible | Yes | Yes | Yes | Yes |
| | | ← (Figure 6.2) → | | (Figure 6.8) | ← (Figure 6.10) → | | ← (Figure 6.10)→ | |

*Notes*
1 All values are approximate and are for reference only. For exact values consult the manufacturer.
2 All these drives are based on pulse width modulation (PWM) and hence would produce over-voltages at the inverter output and require over-voltage protection for cable lengths of 100 m (typical) and above, depending upon the steepness of the wave (Section 6.13.2).
3 The performance of the drive would also depend upon the accuracy of the motor's mathematical model used for the phasor control.
4 The choice of the type of drive would depend upon the degree of speed regulation required by the process.
[a] Open-loop controls
[b] Closed-loop controls
[c] These drives are normally open loop (sensor-less) without encoder. For higher regulation, it is better to adopt a two-phasor control, such as a field-oriented control (FOC) or a direct torque control (DTC) drive.
[d] Response time without an encoder is sufficiently low. Where response time alone is the prime consideration, the encoder is not necessary.

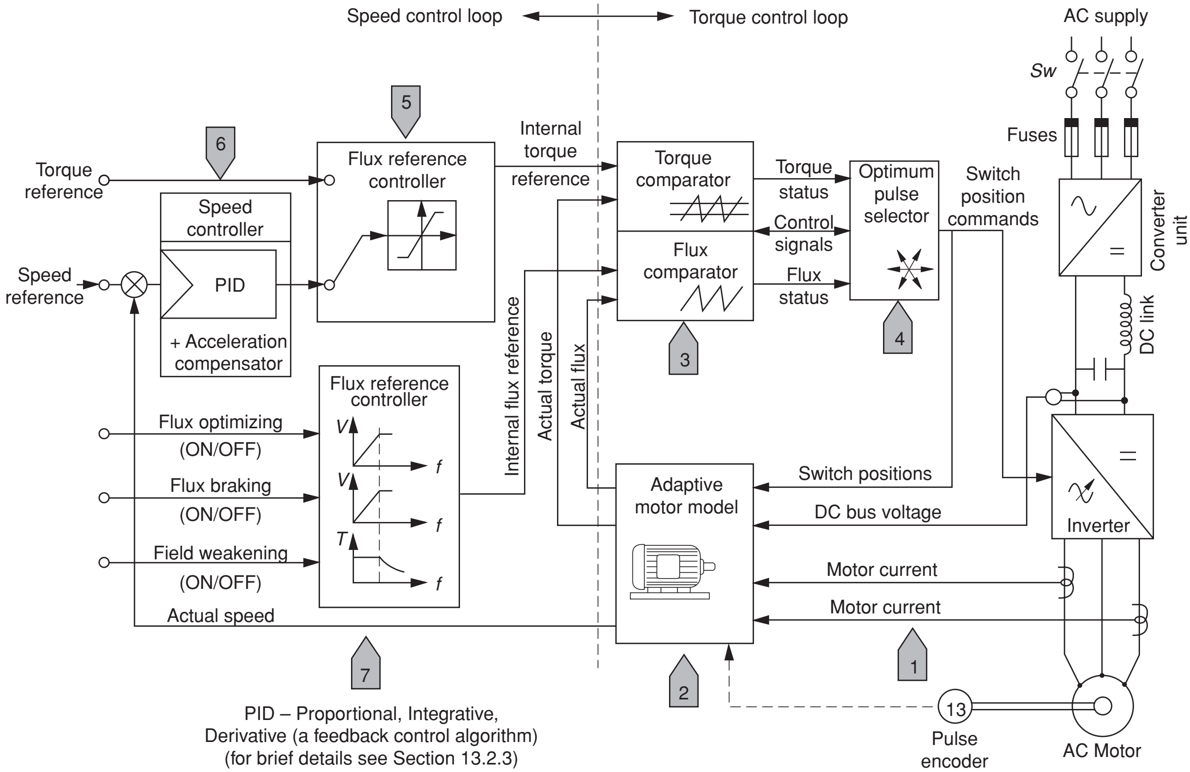



**Figure 6.13** Block diagram of a direct torque control inverter circuit (Source: ABB)

**Section 2**
This is a highly advanced motor model, which is first made to read and store the machine's vital parameters such as $R_1$, $L_1$, saturation coefficients and its moment of inertia during an autocalibration run. The motor is run under a locked rotor condition and the mathematical model is capable of computing its basic characteristics in terms of these parameters or any data that may be of use to actuate the control logistics. The block diagram is drawn in an open-loop condition, which would suffice for most process lines. For still higher accuracy in speed control, an encoder may be introduced into this circuit, as shown by the dotted line in Figure 6.13. The quantities measured in Section 1 are also fed into this section, which are able to compute the actual operating values of $T$ and $\phi$ about every 25 $\mu$s, i.e. around 40 000 times every second. These are the output control signals of this section.

**Section 3**
The above actual operating data are fed into a torque comparator and a flux comparator, which are already supplied with predefined reference data to which the machine is required to adjust. Any errors in these two data (reference and actual) are compared by these comparators at an extremely fast speed as noted above and provides $T$ and $\phi$ error signals to the switching logistics of the inverter unit.

**Section 4**
This determines the switching pattern of the inverter unit, based on the $T$ and $\phi$ error signals, obtained from the torque and flux comparators. Since these signals are obtained at very high speed, the inverter IGBTs are also switched with an equally high speed to provide a quick response and an accurate $T$ and $N$.

## 2 Speed control loop

**Section 5**
This is a torque reference controller which controls the speed control output signal through the required torque reference signal and the d.c. link voltage. Its output torque reference is fed to the torque comparator (Section 3).

**Section 6**
This is the speed controller block that consists of both a PID (proportional integrative derivative, a type of programming) controller and an acceleration compensator. The required speed reference signal is compared with the actual speed signal obtained from the motor model (section 2). The error signal is then fed to both the PID controller and the acceleration compensator. The sum of these two is the output signal that is fed to the torque reference controller (Section 5).

**Section 7**
This is the flux reference controller which provides the absolute value of stator flux to the flux comparator (Section 3). The value of this absolute flux can be varied to fulfil many functional requirements from the inverter unit such as

- Field strengthening – to obtain speed variation below the base speed
- Flux braking – to carry out braking duties
- Field weakening – to obtain speed variation above the base speed.

***Inference***

1 With the application of static drives, induction motors, both, squirrel cage and slip-ring, can be easily controlled to achieve the required characteristics by the application of solid-state technology.
2 With the availability of phasor control technology, by which one can separate out the active and magnetizing components of the motor's stator current and vary them individually, it is now possible to achieve higher dynamic performance and accuracy of speed control in an a.c. machine similar to and even better than a separately excited d.c. machine.
3 With this technology it is now possible to achieve extremely accurate speed control of the order of $\pm 0.01\%$ to $\pm 0.001\%$. To achieve such high accuracy in speed control, closed-loop feedback control systems and microprocessor-based control logistics can be introduced into the inverter control scheme to sense, monitor and control the variable parameters of the motor to very precise limits.
4 A very wide range of speed controls is available through this technology as it is possible to vary frequency on both sides ($\pm$) of the rated frequency.

## 6.5 Use of phasor control for flux braking

It is possible to perform braking duties by the motor by raising the level of magnetization (field strengthening). By raising flux, the speed reduces (Figure 6.7(b)) and the stator current rises. This is an apparent advantage in this kind of a speed control, as the heat generated by the motor during braking appears as thermal energy in the stator rather than in the rotor. Also it is easier to dissipate heat from the stator than from the rotor due to its (stator) bulk and outer cooling surface, which is open to the atmosphere.

## 6.6 Control and feedback devices

The control and feedback circuits are also solid-state devices and offer high reliability and accuracy. The output from these devices can be interfaced with a microprocessor to carry out the required corrections in the system's parameters through the inverter controls. A microprocessor is a semiconductor device and consists of logic circuits in the form of ICs (integrated circuits), capable of performing computing functions and decision making. These capabilities are used in carrying out process corrections by providing the necessary timing and control corrective signals to the switching logistics of the inverter unit.

A microprocessor is capable of solving complex mathematical problems very rapidly and can analyse a system more closely and accurately. It can be fed with a variety of measuring and control algorithmic software that may be necessary for system monitoring and control. It is also capable of performing supervisory and diagnostic functions and carry out historical recording to store information related to process and drive conditions, to facilitate reviews of data for continuous process monitoring, fault analysis, diagnostics, trend analysis, etc.

The control logistics such as PWM or frequency controls are digital circuits and are microprocessor based. They can compare the actual inverter output parameters with pre-set reference parameters and help to implement the required precise adjustments in the system's parameters instantly by providing corrective command signals to the switching circuits of the inverter unit. This in turn adjusts the system variable parameters within the required limits by adjusting $V$ and $f$ as in a simple $V/f$ control or $I_m$ and $f$ as in a flux phasor control (single phasor control), or $I_a$, $I_m$, $\beta$ and $f$ as in a field-oriented control or the torque phasor control as in a direct torque control technique etc. As a result of their accuracy and speed, they are capable of achieving prompt corrections, high reliability, better flexibility and hence a high dynamic performance of the drive.

There are many types of sensors used to feedback the process operating conditions to the switching logistics of an inverter unit. They can be in terms of temperature, pressure, volume, flow, time or any activity on which depends the accuracy and quality of the process. Direct sensing devices used commonly for the control of a drive and used frequently in the following text are speed sensors, as noted below.

### 6.6.1 Speed sensors

These are closed-loop sensing devices and are mounted on the machine or a process line. They are able to sense the operating parameters and provide an analogue or digital feedback input to the inverter switching logistics. For example:

1 **Tacho generator (TG)** – This is an analogue voltage feedback device and can provide a speed input to the inverter control circuit. It provides only a low level of speed regulation, typically $\pm 0.4\%$ of the set speed.
2 **Pulse encoder** – This is a digital voltage feedback device and converts an angular movement into electrical high-speed pulses. It provides speed and also the angular position of the rotor with respect to the stator field when required for field-oriented control to the inverter control circuit to achieve the required

speed control. It is a high-accuracy device, and provides accuracy up to ± 0.001% of the set speed. (See the simple feedback control scheme shown in Figure 6.12.)

## 6.7 Evolution of solid-state technology

This field is very vast and a detailed study of the subject is beyond the scope of this handbook. We will limit our discussions to such areas of this subject that relate to the control of a.c. motors and attempt to identify the different solid-state devices that have been developed so far and put to use with their application in the control of a.c. motors. Only the more common circuits and configurations are discussed. The brief discussion of the subject provided here, however, should be adequate to help the reader understand this subject in general terms and to use this knowledge in the field of a.c. motor controls to achieve from a soft start to a very precise speed control and, more importantly, to conserve energy of the machine which is wasted otherwise. For more details of static controllers and their design aspects see the Further Reading (Sr. nos. 2, 4, 5, 8, 11, 12 and 15 onwards) at the end of the chapter or consult the manufacturer. To bring more clarity to the subject of power semiconductor devices we have also discussed their applications in different fields other than motor controls. At appropriate places passing references to a d.c. machine are also provided. To make the discussion more complete different configurations of converter units are also discussed for the control of d.c. motors.

In the past more than five decades solid-state devices such as diodes, transistors and thyristors have attained a remarkable status and application in the field of electronic power engineering. Diodes and thyristors were introduced in the late 1950s, while the basic transistor (BJT – bipolar junction triode) was introduced in 1948. In India they appeared much later (thyristors were introduced in the 1970s and power transistors in the 1980s). This technology is now extensively applied to convert a fixed a.c. power supply system to a variable a.c. supply system, which in turn is utilized to perform a required variable duty of a fixed power system or a machine. They are all semiconductor switching devices and constitute two basic families, one of transistors and the other of thyristors and these two resulting into their hybrids also like MCTs and IGCTs (MOS – controlled thyristors and integrated gate commutated thyristors). The more prevalent of them are discussed here to give readers an idea of the use of this technology in today's domestic and integrated applications, power generation, transmission, distribution and their controls etc.

More emphasis is provided on the control of induction motors. Research and development in this field is a continuous process and is being carried out by agencies and leading manufacturers. This aims to advance and optimize the utility of such devices by improving their current-handling capability and making them suitable for higher system voltages, switching speeds etc. There may be more advanced versions available in course of time and readers are advised to contact the leading manufacturers for details on the latest technology.

Diodes are purely static power switching devices and are used extensively with thyristor and transistor power schemes. Transistors are relatively cheaper and easy to handle compared to thyristors. The latter are more expensive and more complex as noted below. With the use of these devices an induction motor can be employed to perform variable duties through its stepless speed control logistics by close monitoring of load requirements during a particular process or while performing a specific duty cycle. The controls are assisted by microprocessor-based, open- or closed-loop control techniques, which can sense and monitor many variables such as speed, flow of material, temperature, pressure or parameters important for a process or a duty cycle. With these techniques, it is possible to achieve any level of automation. Open-loop systems are used where high accuracy of controls and feedback is not so important and closed-loop where a high degree of accuracy of control is essential. With solid-state technology it is now possible to utilize a conventional machine to perform a variable duty. The advanced versions of these devices are capable to handle large powers and with their series parallel combinations very very large powers. They can now handle all kinds of power requirements and their controls. Like HV d.c. power transmission, reactive power controls, large rectifiers and all sizes and duties of induction motors.

### 6.7.1 Power diodes

These are uni-directional* and uncontrollable† static electronic devices and used as static switches and shown in Figure 6.14. A diode turns ON at the instant it becomes forward biased and OFF when it becomes reverse biased. By connecting them in series parallel combinations, they can be made suitable for any desired voltage and current ratings. So far these devices have been developed up to 6.5 kV, 6.0 kA and even more. Whether it is a transistor scheme or a thyristor scheme, they are used extensively where a forward conduction alone is necessary and the scheme calls for only a simple switching, without any control over the switching operation. They are used extensively in a rectifier circuit to convert a fixed a.c. supply to a fixed d.c. supply.

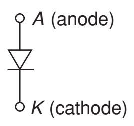



**Figure 6.14** Circuit symbol for a power diode as a switch

* A unidirectional switch is one that can conduct in only one direction and blocks in the reverse direction.

†A controllable switch is one that can be turned ON and OFF by switching a control circuit ON and OFF.

## 6.7.2 The power transistor family

The solid-state technology in the field of transistors in particular has undergone a sea-change, beginning in the 1950s from the basic bipolar junction transistor (BJT) to the more advanced insulated gate bipolar LV transistor (IGBT) by the 1990s, integrated gate commutated thyristors (IGCTs) by the 1997 (a hybrid of transistors and thyristors), MV IGBTs by 1998, symmetrical gate commutated thyristors (SGCTs) by 1999 and latest transistorized power device injection enhanced gate transistors (IEGTs) by 2000. The following are some of the more prominent of the power transistor family that are commonly used in power circuits.

### *Bipolar junction transistors (BJTs)*

These are the basic transistors (triodes) and are illustrated in Figure 6.15. They are uni-directional and controllable and are capable of handling large currents and high voltages and also possess high switching speeds (faster than thyristors). However, they require a high base current due to the high voltage drop across the device, which causes a high loss and dissipation of heat. This adverse feature of their characteristics renders them unsuitable as power switching devices for efficient power conversion. Therefore they are generally used as electronic control devices rather than power devices in electronic control circuits and are not produced in higher ratings.

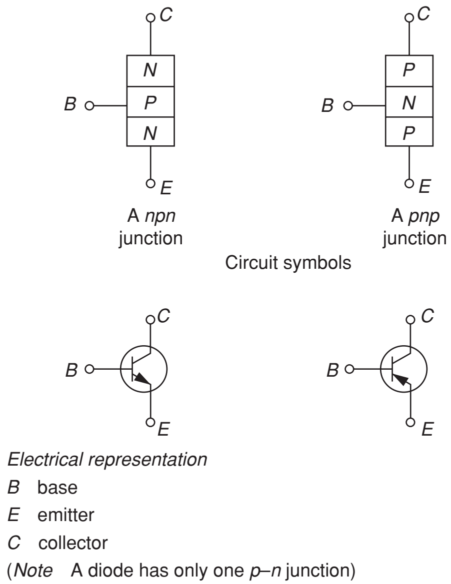



**Figure 6.15** Circuit symbols and electrical representation of a basic triode or power transistor (BJT)

### *Two-junction transistors or power Darlingtons*

These also have three terminals as illustrated in Figure 6.16. They are fabricated of two power transistors and are used as a single transistor and are suitable only for control circuits. They are used to reduce the control current requirement and hence cause lesser heat dissipation, particularly during switching operations.

### *MOSFETs*

These are metal oxide semiconductor field effective transistors and are shown in Figure 6.17. They are capable of switching quickly (but are more sluggish than BJTs) and handle higher switching frequencies. But they can deal with only lower currents and withstand lower voltages and thus possess a low power-handling capability. They are bi-directional and can operate as controlled switches in the forward direction and uncontrolled switches in the reverse direction. MOSFETs are composed of a diode and a BJT or IGBT in anti-parallel. They are voltage controlled devices and require a negligible base current at their control terminals to maintain the ON state and hence are low-loss devices. MOSFETs are used extensively as fully controllable power switches, where they are required to handle only low powers. The normal trend is to use them only as control devices.

The power BJTs and power MOSFETs have provided two very useful static switching devices in the field of transistor technology. But while the former can handle larger powers, they dissipate high heat, the latter poses a limitation in handling large powers. As a result of these shortcomings, they are used mostly as control circuit switching devices. Such limitations were overcome by yet another development in the 1990s, in the form of an IGBT.

### *Insulated gate bipolar transistors (IGBTs)*

These are unidirectional transistors and have an insulated gate (G) instead of base (B) as in a bipolar transistor (BJT) and are represented in Figure 6.18. They are a

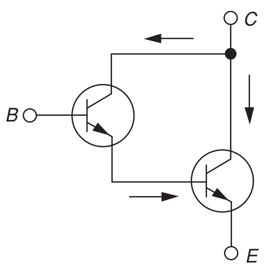



**Figure 6.16** Circuit symbol for a two-junction transistor or power Darlington

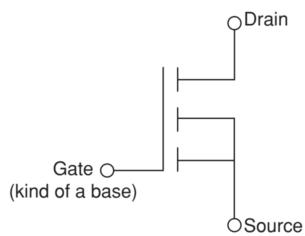



**Figure 6.17** Circuit symbol for a power MOSFET

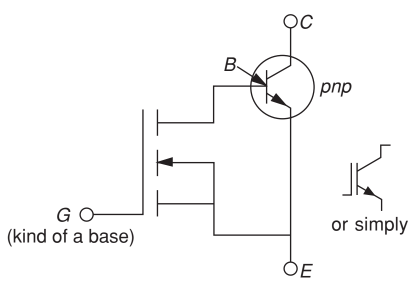



**Figure 6.18** Circuit symbol for an IGBT

hybrid of a pnp bipolar transistor which is connected to a power MOSFET like a two-junction transistor (Power Darlington, Figure. 6.16). A positive voltage between the gate and the emitter, switches ON the MOSFET and provides a low resistance effect between the base and the collector of pnp bipolar transistor and switches this ON as well. The combination of two transistors offers an insulated gate that requires a low base current which makes it a low-loss device. When the voltage between the gate and the emitter is reduced to zero, the MOSFET switches OFF and cuts off the base current to the bipolar transistor and switches OFF that as well. With slight modification in construction and upgradation of the bipolar and MOSFET transistors, it is possible to produce a low-loss IGBT, suitable for fast switching, handling large currents and withstanding higher voltages. The turn-on and turn-off delay is about 1 μs only.

The dramatic change and advancement of this hybrid combination in a bipolar transistor has greatly enhanced the application of power transistors in the field of power conversion, variable speed drives and energy conservation. It possesses the qualities of the power bipolar transistor (BJT) as well as the power MOSFET. Like a power MOSFET, it too is a voltage controlled switching device that permits fast switching and gate voltage control. And as a bipolar transistor it allows a large power handling capability. The switching speed of IGBT is also higher than that of a bipolar transistor. It thus provides an efficient power conversion device.

### *IGBTs for LV systems*

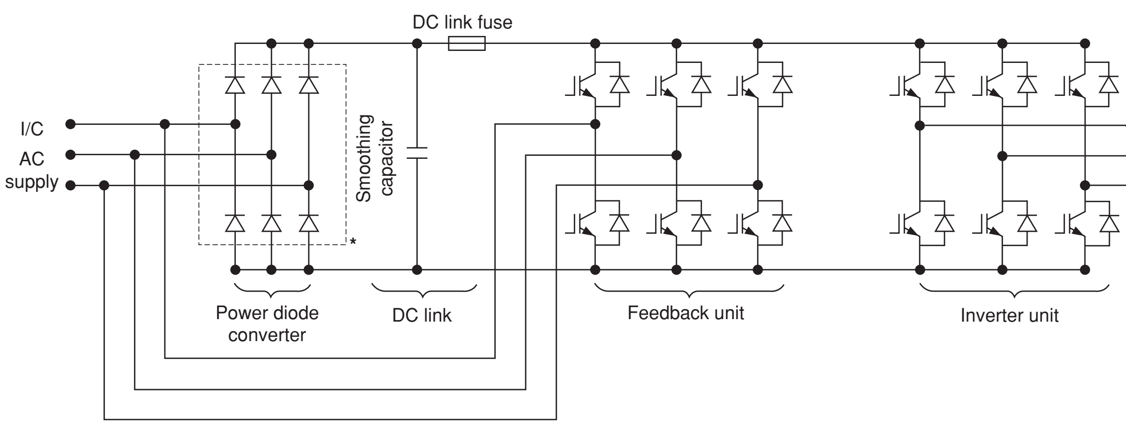
During 1990s they had revolutionized industrial power conversion in the low voltages. Single IGBT inverters on low voltages (690–1000V) can be economically designed up to about 1.4 kA or about 1 MVA with forced air-cooling and up to about 1.5 MVA with liquid (ethyl-glycol) cooling. For still higher ratings, they may be arranged in parallel to obtain higher currents. But parallel switchings may pose some limitations in simultaneous switchings of the devices and affect their reliability, though it has been overcome to a great extent. Nevertheless low voltage IGBTs up to about 1.5 MVA are seen to be more economical and a unanimous choice for all standard applications.

They are used extensively in a voltage source inverter (VSI) circuit to convert a fixed d.c. supply to a variable a.c. supply for motor drives. Since they are more expensive compared to power diodes, they are usually not used in a rectifier circuit where power diodes are mostly used.

However, when the power is to be fed back to the source of supply, they are used in the rectifier circuit also to adjust $V$ and $f$ of the feedback supply with that of the source. Such as during energy conservation when being used as a motor drive or during power conversion when being used in a wind generator (Section 7.21).

A typical inverter circuit using IGBTs is illustrated in Figure 6.19. In addition to being suitable for high switching frequencies (typically 2–30 kHz), these transistors generate very low harmonics of the order of 1% or less of the system voltage and virtually retain the sinusoidal waveform of the motor currents. The motor current containing lesser harmonics also causes lesser heating of the motor windings. In most cases it eliminates the use of costly filter circuits making them yet more economical. It also causes lesser pulsation in torque and low motor noise. The motor also runs smoother, even at low speeds. Now $V$ and $f$ both can be varied through this single device and a fixed voltage power diode bridge rectifier is sufficient to obtain fixed d.c. voltage, rather than using a phase-controlled thyristor rectifier to obtain a variable d.c. voltage.

IGBTs themselves do not generate switching surges but when switched frequently such as during PWM operation, the output supply will contain moderate switching surges that may be detrimental to the motor windings as discussed in Section 6.13.2. Depending upon the type of installation, a surge protection, in the form of d$v$/d$t$ protection through a snubber circuit (Figure 6.37) or a d$i$/d$t$ protection through a choke or a combination of




* Uncontrolled line side diode bridge rectifier. When a variable d.c. is required, it can be replaced by thyristors.

**Figure 6.19** A typical inverter circuit using IGBTs, also showing a feedback unit both to attain a yet more accurate sinusoidal waveform (Section 6.13.2) may become essential at the output (motor side) of the inverter with such drives. Particularly when the cable length from the drive to the motor is long or when the motor is rather old and may not be possessing a sound dielectric strength. More details on this aspect are discussed in Section 6.13.

### *IGBTs for HV Systems*

Initially it was attempted to use LV IGBTs in series to raise their operating voltages. But experience with series connected IGBTs for HV applications has not been encouraging because of some limitations such as,

– Problem in simultaneous switchings of devices affecting their reliability
– Higher losses during ON-state and turn ON transition
– Very high turn off d*i*/d*t*
– Series parallel combination requires more space
– It makes them expensive at higher operating voltages
– More number of devices also lead to lesser reliability.

These limitations of IGBTs on medium voltages shifted thrust on thyristor based devices that culminated in IGCTs and SGCTs as discussed later. But work on IGBTs continued and with gradual and consistent development in their design, it has now been possible to develop HV IGBTs also with much larger ratings such as up to 4.5 kV, 1.2 kA as noted in Table 6.2a. They also possess the following features

– They are compact and possess high efficiency
– Can perform at high switching frequencies
– Possess high blocking voltage and snubberless operation with simple firing circuit
– Switching surges are low
– They require only about 5 A to switch it ON or OFF.
– Intermediate control is possible up to any level between ON and OFF.
– They can control the current on their own. The emitter current being a function of gate emitter voltage and temperature. During a fault condition on the inverter output side the voltage drop across the emitter and collector rises but the voltage across the gate and emitter is controlled by the gate drive circuit and hence the fault current is limited automatically.
– Transistor family becoming capable to handle such large powers on LV and MV systems is surely an intrusion into the domain of GTOs and thyristors (SCRs).

*Note*
They are now more reliable and simple but at still high powers when using a number of them in series and parallel they may have switching problems as noted above.

***Applications***: In inverter circuits for
LV IGBTs – Sugar centrifuges, conveyor belts, pumps, fans, compressors, ID fans and cement plants.
HV IGBTs – Ship propulsion drives, frequency converters or PWM converters with constant d.c. voltage, metal industries (earlier cyclo converters or load commutated inverters (LCIs) or GTO inverters were used). Descaling pumps in steel plants. Cold rolling mills.

### *Injection enhanced gate transistors (IEGTs)*

IEGT is yet another addition in the family of transistors. It is a high power handling semiconductor device making use of transistor technology and also retaining its fast switching capability. It can be operated at high frequency (several kHz) and provide smooth control of machines. It is compact, has low losses and is more economical. Rating is provided in Table 6.2a.

There are yet more switching devices in the field of

**Table 6.2a** Ratings of high capacity power devices developed so far

| *Device* | *Rating* | |
|---|---|---|
| | *Voltage kV* | *Current kA* |
| Power Diodes | | |
| General use | 2.8 | 3.5 |
| Fast switching | 6.0 | 3.0 |
| SCRs | | |
| General use | 12.0 | 1.5 |
| Fast switching | 1.2 | 1.5 |
| Light triggered | 8.0 | 3.6 |
| GTOs | 6.0 | 6.0 |
| IGBTs | | |
| LV | 1.0 | 1.2 |
| HV | 4.5 | 1.2 |
| MCTs | 3.3 | 1.2 |
| IGCTs | 6.0* | 1.2 |
| | 6.5* | 6.0 |
| SGCTs | 6.0 | 5.0 |
| IEGTs | 4.5 | 4.0 |

Based on Toshiba, GE, Alstom & Mitsubishi Electric technical reports.

* Under development

*Notes*
1. Ratings noted above are only for a general reference and may vary from manufacturer to manufacturer.
2. Ratings up to 3–4 MVA usually refer to with air cooling and would rise with liquid cooling.
3. These are all high power handling devices. There being no rule of thumb to decide the use of a particular device for a particular application. It can depend upon the inventions and the manufacturing practices of a manufacturer, availability of vital components and most importantly cost consideration. Brief details provided in the present text, however, will provide a fairly good idea to the reader or a user about the applications of these devices.
4. With the fast changing semiconductor technology and transistor or transistor-thyristor based hybrid switching devices gradually becoming state-of-the-art large power handling devices, thyristors (SCRs) and GTOs are gradually losing their hold.

semiconductor devices as developed by different manufacturers, such as

- MOS controlled gate turn-off thyristors (MCTs), a hybrid of MOSFETs and gate turn-off thyristors
- Integrated gate commutated thyristors (IGCTs), a hybrid of bipolar transistors (BJTs) and gate turn-off thyristors and
- Symmetrical gate commutated thyristors (SGCTs).

All these devices are hybrid of transistors and thyristors. As they are capable of handling large powers, we have discussed them with thyristors in the next section.

Continuous research is a buzzword with semiconductor devices to further enhance their ratings, switching capabilities, efficiency of operation and reduction in losses. Different manufacturers may assign different trade names to identify their products depending upon the device and the technique used. In the brief text provided here we have discussed only the basic devices and their applications and it is expected that it will suffice the reader to know the developments in the field of electronics and their applications in large to very large power handling.

### 6.7.3 The thyristor family

The thyristor is a semiconductor device made of germanium or silicon wafers and comprises three or more junctions, which can be switched from the OFF state to the ON state or vice versa. Basically it is a *pnpn* junction, as shown in Figure 6.20(a) and can be considered as composed of two transistors with *npn* and *pnp* junctions, as illustrated in Figure 6.20(b). It does not turn ON when it is forward biased, unlike a diode, unless there is a gate firing pulse. Thyristors are forced commutated (a technique for switching OFF) and hence call for a complex circuitry, more so in a 3-$\phi$ system, where six of them have to operate simultaneously. The control device has to be very accurate to trigger all the thyristor devices at the same instant. A slight error in the firing angle may cause a short-circuit, whereas a transistor can be switched OFF simply by removing the base signal. Thyristors are therefore also known as phase-controlled rectifiers. The phase angle delay is known as the firing angle of the switching element. Here we have denoted this angle by $\alpha$. For diodes $\alpha = 0$, while for thyristors it can be adjusted as illustrated in Figure 6.23. We will not go into more detail on the construction of this device and will limit our discussion only to its application in a power system.

They are high power semiconductor devices compared to transistorised devices. The device constitutes a large family, but only the more prevalent of them are discussed here, i.e.

- Silicon-controlled rectifiers (SCRs). These are basically thyristors and unless specified, a thyristor will mean an SCR.
- Triacs
- Gate turn-off switches (GTOs)
- MOS-controlled thyristors (MCTs)
- Integrated gate commutated thyristors (IGCTs) and
- Symmetrical gate commutated thyristors (SGCTs)

#### *Silicon-controlled rectifiers (SCRs)*

The most popular of the thyristor family is the SCR, which was first developed in 1957 by General Electric, USA. The SCR is similar in construction to that of a junction diode, except that it has three junctions instead of one, and a gate to control the flow of power. The SCR is commonly represented as shown in Figure 6.20(a) and can be regarded as a semiconductor switch, similar to a toggle switch. An SCR is uni-directional and conducts in one direction only and can also be termed a reverse blocking triode thyristor (a symmetrical device). When anode A is positive with respect to K, it is in the conducting mode and is termed forward biased. The *V–I* characteristics are similar to those of a semiconductor diode, as shown in Figure 6.21. When K is positive with respect to A, it is in the nonconducting mode and conducts a very low leakage current. In this mode it is termed reverse biased. In this condition, when the reverse voltage is raised a state is reached when the low-leakage reverse current increases rapidly as a result of the dielectric breakdown. This stage is called the reverse avalanche region (Figure 6.21) and may destroy the device.

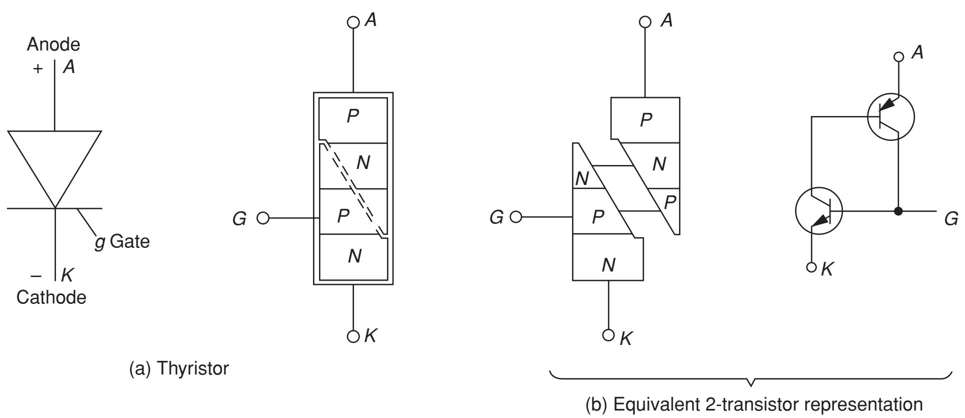



**Figure 6.20** A basic thyristor [silicon controlled rectifier (SCR)]

When a load and a power source is connected across the anode and the cathode of the SCR, there will be no conduction and no current will flow, even when the anode is made positive with respect to the cathode unless the gate is also made forward biased with the application of a positive potential at the gate. After the conduction commences, the gate potential can be removed and the device will continue to conduct. It is the gate signal that plays the most vital role in achieving the desired variation in the voltage. The main power connections to the device are made to its terminals A and K and a turn-on signal is applied between the gate and K. An SCR can easily provide a variable voltage source by varying its firing angle. In view of its simplicity, it is the most commonly used thyristor in a phase-controlled rectifier unit (converter). Gate control is now simple, as it is connected on the a.c. or the line side, which provides it with a natural commutation. The thyristor gets switched OFF at every current zero. This may therefore also be termed a line commutated rectifier.

The use of SCRs in an inverter circuit is intricate because of the absence of a natural commutation. Now only a forced commutation is possible, as it is connected to a d.c. source which provides no current zeros and hence facilitates no natural commutation. A forced commutation calls for a separate switching circuit, which is cumbersome, besides adding to the cost. As a result of this feature, they are also called forced commutated thyristors.

### *Triacs*

Unlike an SCR, which is unidirectional making them suitable only for d.c., a triac is a bidirectional thyristor switch and conducts in both directions. It can be considered as composed of two SCRs, connected back to back with a single gate, as shown in Figure 6.22(a). Since the thyristor now conducts in both directions there is no positive (anode) or negative (cathode) terminals. They now become suitable for a.c. supply. But in view of their very low power handling capability they are no longer used for heavy duty applications. Their application is limited to only control of low power devices such as small single phase motors, fans and domestic appliances.

### *Reverse conducting thyristors*

The triacs inherit some limitations in handling frequencies higher than normal. In such cases, they can be simulated by using two SCRs in inverse parallel combinations as illustrated in Figure 6.22(b). Now it is known as a reverse conducting thyristor. An SCR has no frequency limitations at least up to ten times the normal. The required voltage and current ratings are obtained by series-parallel connections of more than one thyristor unit.

### *Gate turn-off switches (GTOs)*

This is a development of 1980s. The gate can only turn the thyristor ON but it cannot turn it OFF (commutate). Switching OFF can be accomplished only by reducing the conducting current to less than the thyristor's holding current which may be up to 1/3rd the load current. To achieve such a high gate current in amplitude and $di/dt$ several high power MOSFETs are required in parallel in the firing circuit and the circuit must possess an extremely low impedance. A device that allows the gate to switch OFF is called the gate turn-off switch (GTO) or gate control switch (GCS). The GTO turns it OFF by firing (applying) a negative potential between the gate and the cathode as shown in the modified representation of an SCR, Figure 6.22(c). It therefore calls for a sophisticated firing circuit. Since they allow only ON or OFF conditions, intermediate controls are not possible.

Once they are ON, external resistance or reactance alone can limit the current through GTOs. It is the most commonly used device in a thyristor inverter circuit. They are rugged and robust and can handle high voltages and currents (typically up to 6kV and 6kA). Their switching speed is very high (up to 10–100 ns (n is $10^9$)) but lack high frequency switching (<10 kHz) and require a snubber

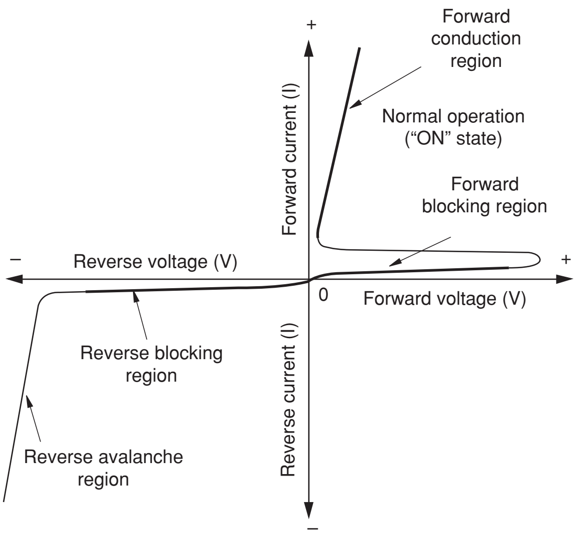



**Figure 6.21** *V-I* characteristics of a thyristor (SCR) without a gate voltage

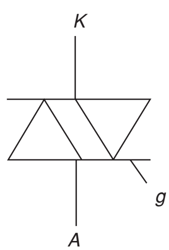



**Figure 6.22(a)** Schematic representation of a triac

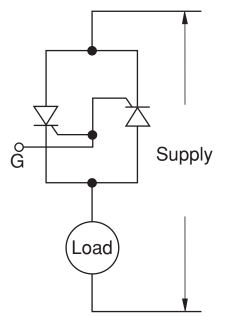



**Figure 6.22(b)** Use of two SCRs in inverse parallel combination to simulate a triac (reverse-conducting thyristors)

during turn-off. A short-circuit on the inverter output side can be quite severe and calls for an effective short-circuit protection.

***Applications*** – Suitable for handling very large powers where IGBTs, IGCTs and SGCTs may pose a limitation.

### *MOS-controlled thyristors (MCTs)*

They consist of two MOSFETs and a GTO. MOSFETs are used to have full control over the thyristor as shown in Figure 6.22 (d). Upper MOSFET is used to trigger the thyristor that results in an extremely low gate current requirement. A positive gate voltage triggers the device to switch on while a negative one forces it to turn off. Zero gate voltage allows it to maintain status-quo. They are used for MV systems and have similar utilities as of IGCTs.

### *Integrated gate commutated thyristors (IGCTs)*

They are now a reality and can handle very large powers. They can be considered an improvised version of GTOs with integrated gate driver utilizing their excellent turn off advantages. They are represented as shown in Figure 6.22(e). Their power handling capabilities achieved so far are noted in Table 6.2a. Leading manufacturers worldwide are gradually using this device in their new large motor drives and other high power handling applications. They have been specially developed with a focus on a 2 level, 3 level and multiple level voltage source inverters (VSIs) with d.c. bus capacitor. For current source inverters (CSIs) a reverse blocking element is required for which symmetrical GCTs (SGCTs) have been developed and are discussed next.

The IGCTs combine the excellent forward characteristics of thyristors and switching performance of bipolar transistors. As they use thyristor technology they are rugged and robust. By means of a "hard-drive" gate control with unity gain turn-off, the element changes directly from the thyristor mode to the transistor mode during turn-off. This allows operation without any snubber and reduces the total switching losses to only 50% that of GTOs at load and nearly zero at no-load. They can have a switching frequency up to 2–3 kHz and gate power 60–70% that of GTOs. They also (like IGBTs) generate only low harmonics of the order of 1% or less of the system voltage. IGCTs are available as disc-type elements which are reverse conducting type (asymmetrical). The reverse conducting IGCTs have an anti-parallel freewheeling diode, which is required for most switching operations and can be a separate device or integrated into the wafer. This reverse conducting element adds to significant losses to the complete power structure and limits the maximum permissible d$i$/d$t$ capability during the turn ON transition.

Gate turn – ON or OFF

- An IGCT requires almost the same current at the gate as the momentary load current. To achieve such a high gate current in amplitude and d$i$/d$t$ several high powered MOSFETs can be required in parallel in the firing circuit and the circuit must have an extremely low impedance.
- They are capable of high switching frequency, compared to GTOs and SCRs and generate high switching surges.
- They have high blocking voltage and snubberless operation.

Since IGCTs are not provided with insulation to ground, the individual semiconductor switches are mounted onto separate heat sinks which are insulated if necessary. The gate driver required is part of the IGCT. Because of ease of operation, switching speed, reduced losses and high power handling capability they too like IGBTs are an intrusion into the domain of thyristors (SCRs) and GTOs. IGCTs are however, economical only for high power requirements 1MVA and more. They permit construction of air-cooled solutions for 2-level inverters

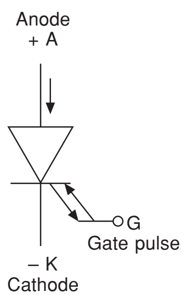



**Figure 6.22(c)** Representation of a gate turn-off-switch (GTO)

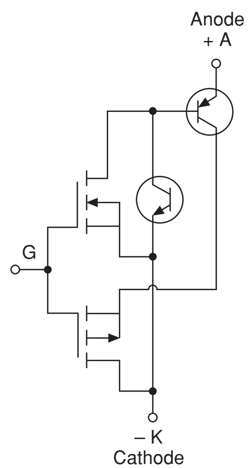



**Figure 6.22(d)** Circuit symbol for a MOS controlled thyristor (MCT)

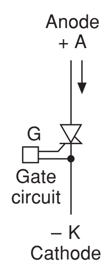



**Figure 6.22(e)** Representation of an integrated gate commutated thyristor

up to about 5 MVA and 3-level inverters up to about 10 MVA without a need to make parallel circuits. With liquid cooling, these ratings can even be doubled. A typical inverter circuit using IGCTs is shown in Figure 6.22(f) which is similar to IGBTs circuit except its inverter unit that now replaces IGBTs with IGCTs. Figure 6.22(g) shows IGCT wafers and an IGCT drive unit in its housing.

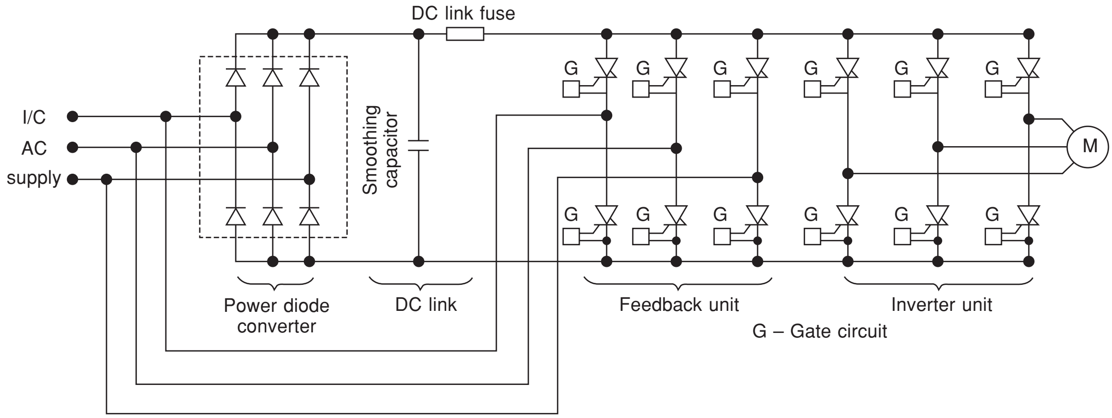



**Figure 6.22(f)** A typical three-phase inverter circuit using IGCTs. Also showing a feedback unit

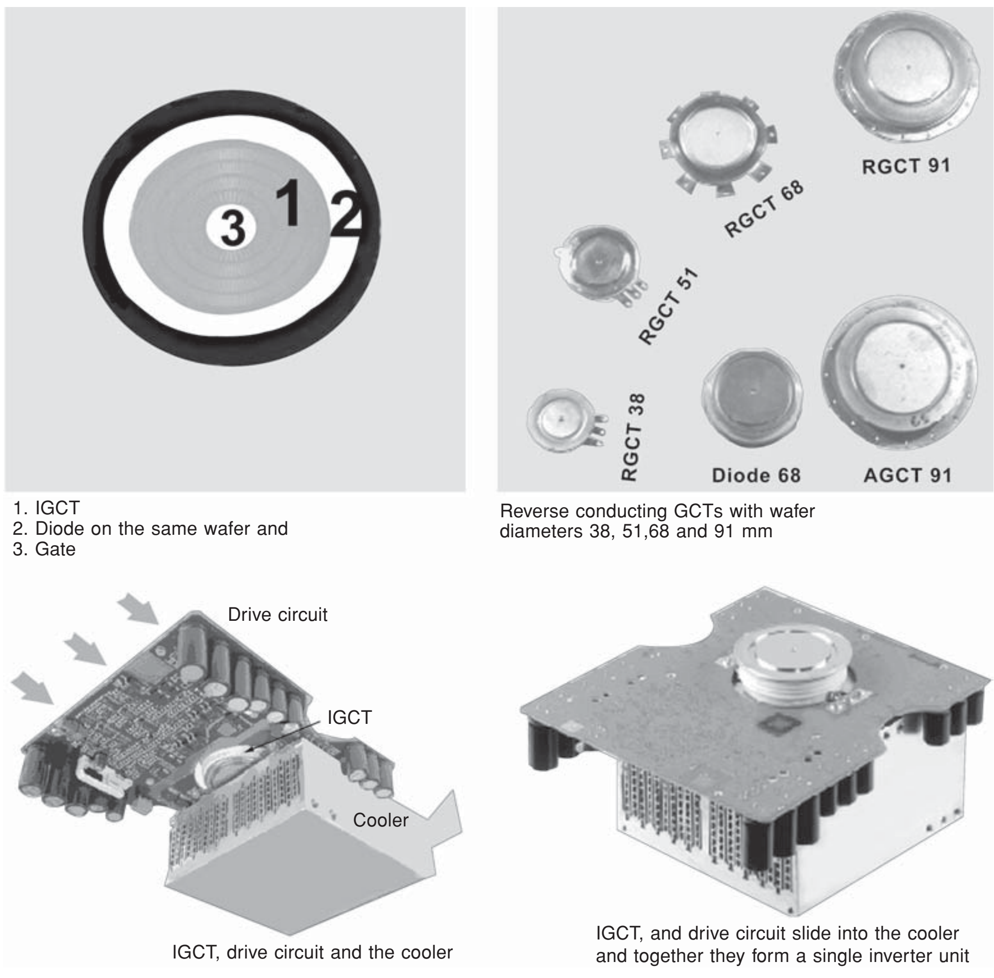



**Figure 6.22(g)** A 51 mm IGCT disc 6 KV, 520 A (Courtesy: Alstom)

***Applications***

- Converter and inverter circuits
- Traction
- Static var compensators
- Large power handling
- Motor drives for different applications

***Symmetrical gate commutated thyristors (SGCTs)***

They are an advance version of IGCTs and possess reverse blocking capability (making them symmetrical devices) and have no free wheeling diode (unlike in IGCTs that have free wheeling diode and are reverse conducting). In other words they are modified gate turn-off thyristors (GTOs) with an integral gate drive. Integrated gate drive improves their turn-off capability and they can offer higher switching frequencies and double sided cooling. Integrated gate driver and 'hard-drive' gate control make the device to switch off faster than any GTO directly from thyristor mode of the ON static operation to the transistor mode during turn-off transition. Symmetrical wafer processing and creating two blocking junctions on the same silicon wafer and series connection of a diode with an asymmetrical turn-off device can be an IGBT or an IGCT. IGCTs having high blocking voltages and low losses is a more preferred choice. Positioning the gate drive close to GTO makes it a low inductance path providing higher efficiency and uniform gating to the device (SGCT).

Now it can handle fluctuating voltages and currents more easily than a GTO while switching ON and OFF during gating. SGCT blocks voltages in both directions unlike IGCTs that are capable for only forward voltage blocking. The bi-directional voltage blocking characteristic of SGCTs make them most useful in pulse width modulated current source inverters (PWM-CSIs) where current is uni-directional while the voltage can assume both polarities making them inherently regenerative drives.

Because SGCTs are capable of blocking voltage in both directions, many components required for IGCTs or IGBTs control gating and current flow can be reduced in size even eliminated from SGCT drives. Figure 6.22(h) shows SGCTs mounted on a circuit board. For more details consult the manufacturer or see Further Reading.

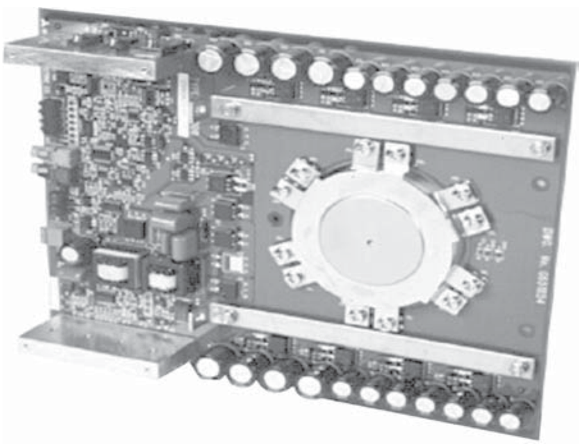


**Figure 6.22(h)** SGCT and gate components mounted on the same circuit board for quick replacement (Source: Rockwell Automation and Mitsubishi (Ref. 22))

They are used in inverter circuits. Brief description of PWM-CSI a.c. drives using these devices is provided in Section 6.9.4. Figure 6.29 illustrates the scheme and Figure 6.22(i) the modules of PWM-CSI a.c. drive. The inverter has current source characteristics at the d.c. terminals using a reactor and voltage source characteristics at the a.c. output terminals using a capacitor. Using sensorless direct phasor control the drive can measure motor flux rather than only making an estimate of it and hence perform better than *V/f*. This also enables the motor torque to be changed promptly similar to a d.c. machine (Figure 6.51).

There are six SGCT switches for 6.6 kV drive. While operating they interface with the a.c. and the d.c. sides. They are operated so as to avoid an open circuit on the d.c. link or a short-circuit on the a.c. side. That means at any instant there are only two switches conducting, one in the top half of the bridge and one in the bottom half. Figure 6.22(j) illustrates a few waveforms,

i. shows the uni-directional current waveform for a typical seven pulse switching pattern
ii. shows the blocked bi-directional voltage waveform
iii. shows the inverter output current and
iv. shows the line to line motor voltage and current

***Applications***

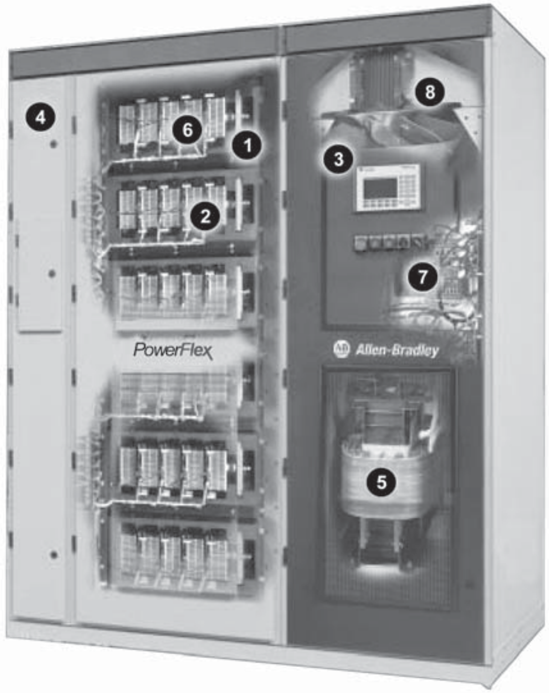
They are high capacity devices usually over 1 MVA and are used for both rectifier (converter) and inverter (VSI and CSI) applications. Figure 6.22(i) shows a typical 6.5 kV SGCT drive.

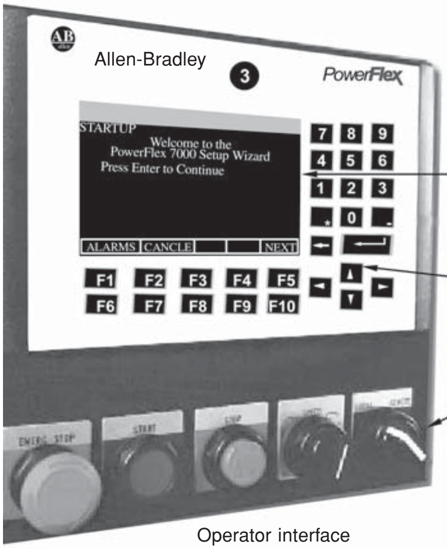
With the availability of IGBTs, IGCTs and SGCTs the three very large capacity semiconductor power switches, there is no bar to applying them to any size of power handling requirement like, static var generators or compensators in reactive power management, d.c. power transmission, large drives and railway traction etc. They are an ideal solution for applications demanding greater stability under fluctuating loads.

### 6.7.4 Applications of power semiconductor devices

Many manufacturers with their extensive R&D back-up are inventing new products every day with better properties, low losses and higher power handling capabilities and introducing them in the market. The semiconductor switching devices considered above are for general reference only and may vary from manufacturer to manufacturer, depending upon their inventions, manufacturing practice and technology in vogue. All these devices are intelligent electronic devices (IEDs) and with the use of telemetry softwares (communication protocols) and media (Section 24.11.5) can transmit data to a remote control station for a total automation of a power system or a process line.

Major applications of these devices have been in the field of soft starting and speed control of a.c./d.c. machines and their active power management (APM) (feeding the




Operator interface

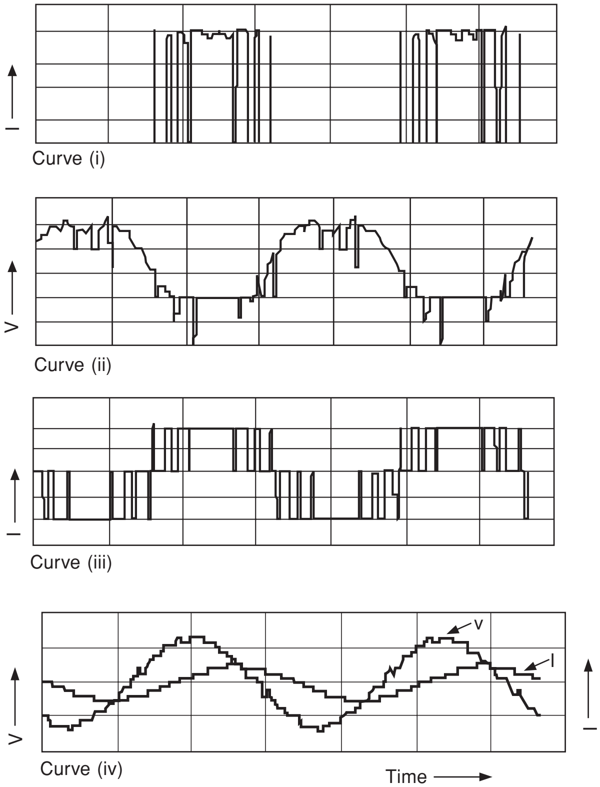
**Figure 6.22(i)** Modules of typical PWM – CSI based a.c. drive (Power Flex 7000) using SGCTs (Courtesy: Allen-Bradley)

surplus power back to the system). These devices have also found their utility in many other industrial and large power handling applications. A few major power applications utilizing these devices are noted below, for the benefit of readers and making the subject a little more lucid and comprehensive.

## 1. HVDC Transmission

It is economical to transmit large powers over short distances ($\simeq$ 50 km) to long distances (>800 km) on d.c. rather than a.c. D.C. causes no skin effect (Section 28.7) and transmits power at unity p.f.

Large power transmission on d.c. with the application of semiconductor devices is no limitation in view of technological advances and high power handling capabilities achieved in these devices discussed already. At the point of transmission a.c. is first converted to fixed d.c which can be up to 500 kV or more. It is then transmitted through a monopolar (single conductor) or bipolar (double conductor) transmission system to the receiving station where it is inverted back to a.c. through




Curve (i) – Device current  Curve (iii) – Inverter output current
Curve (ii) – Device voltage  Curve (iv) – Motor current and voltage

**Figure 6.22(j)** Typical waveforms for a CSI (Source : Rockwell Automation and Mitsubishi (Ref. 22))

an inverter circuit. In case of monopolar system the low voltage return path can be via cables, ground or sea. Basic converter and inverter circuits remain the same as shown in Figure 6.26(a).

D.C. transmission is economical because of reduced cost of transmission due to fewer numbers of conductors. Over long distances the savings can even outdo the cost of conversion in the very first instance and effect huge recurring savings in the long run because of lesser line losses due to,

– Unity p.f. At higher p.fs. the line losses ($I^2R$) are low for the same power transmitted.
– The effective resistance of the conductor does not increase in absence of skin effect. The line losses are therefore further reduced.
– It is estimated that a d.c. transmission system can transmit about threefold more power at the same cost as an a.c. transmission system.

To cope with the ever rising power generation and transmission systems to cater for the constantly rising power demand the world over for industries, services and households, it is becoming a common practice to adapt to d.c. transmission to save on space and cost.

– Back to back stations: These are a.c.-d.c.-a.c. conversion stations and are used to inter-connect two a.c. power transmission or HV distribution systems having different voltages and/or frequencies. A feature of static technology not possible by other means. These stations convert the different voltages and frequencies of different systems to a fixed d.c. voltage and then invert it back to the desired a.c. voltage and frequency. Figure 6.22(k) shows a valve stack unit consisting of GTOs for HVDC back to back stations. A simplified single line diagram is shown in Figure 6.22(k1).

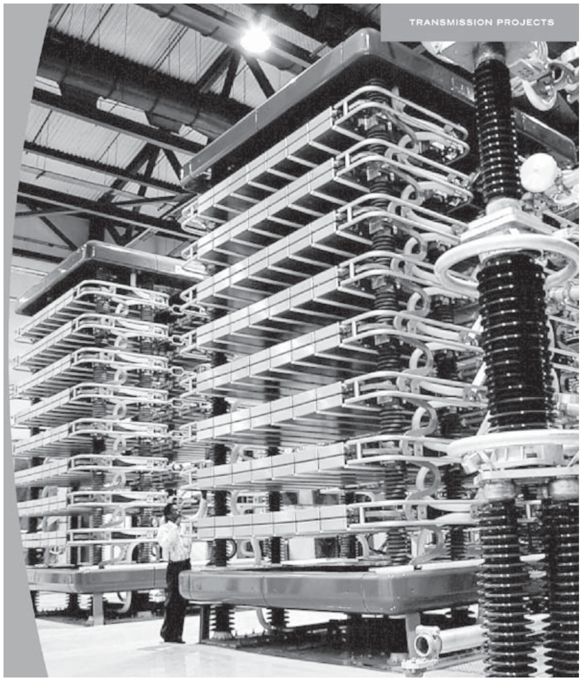



**Figure 6.22(k)** Picture of liquid cooled valve stack consisting of GTOs for HVDC back to back station 2500MW (Courtesy: Alstom)

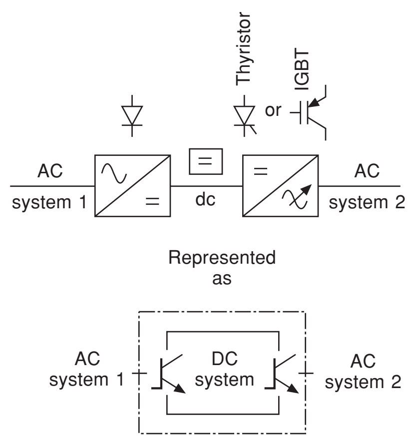



**Figure 6.22(k1)** Back to back stations interconnecting two ac power systems with different voltages and frequencies

– This arrangement can also be extended to a large grid inter-connecting many transmission and distribution systems in form of a d.c. grid as shown in Figure 6.22(k2). Some power systems may have variations in their voltages and/or frequencies and some may be of different voltages altogether. All these different voltages and frequencies are converted to a fixed d.c. voltage and then inverted back to the desired a.c. voltage and frequency for each transmission or HV distribution.
– Such an arrangement can be called a multi-terminal a.c.-d.c.-a.c. configuration. If the grid is feeding some d.c. transmission lines, the d.c. voltage is inverted back to the required a.c. voltage and frequency at the receiving end.

Basic circuit configuration – similar to Figure 6.26(a)
Devices used – SCRs and GTOs

## 2. Static frequency converters (SFCs)

**(i)** ***Conversion of wind power***

Windmills operate at varying speeds because of frequently changing wind pressure. Varying speed means varying frequency and voltage. It is therefore essential to convert these variable parameters to a constant source of supply before supplying it to a grid. This is possible through frequency converters using IGBTs or other devices discussed already (earlier power thyristors were used). The use of frequency converters also helps in achieving low cut-in wind speed and increasing operating hours. For details on windmills see Section 7.21. The wind power generation in practice is usually in the range of 50 kW-6 MW, 0.4–6.6 kV. Other applications of frequency converters in power handling can be mini and micro-hydro power generations.

***Circuit configurations:***

(a) In case of wound rotor generators it will be in the form of a slip recovery system as shown in Figure 6.47. Stator connected to the fixed supply side (grid) and rotor having IGBT converter / IGBT voltage source inverter (VSI) also connected to the same grid to feed the wind-power back to the grid.
Applicable Quadrants (Table 6.4) – 1 and 4 feeding wind-power back to the grid.
(b) In case of cage rotor generators it will be similar to Figure 6.31 but without the additional inverter unit used for feedback as this duty shall be performed by the other inverter unit.

**(ii)** ***Switching of large machines***

Frequency converters can also be used as soft starting devices for large to very large machines 1 MW to 10s of MW by ensuring optimum matching between the SFC and the complete system start-up. Since they are only soft start-up devices they are in the circuit only during start-up period and then removed. They can also be used for two or more machines at a time. One such scheme is shown in Figure 6.22(l) for start-up of a gas turbine or combined cycle (steam and gas turbines together) by running the generator as a motor (as prime mover). Figure 6.22(m) shows an SFCs general arrangement in a modular form.

**(iii)** ***Speed control of machines***

SFCs can also be used for speed control of machines not requiring very accurate speed controls.

***Circuit configuration:***
Load commutated inverter (LCI)
IGCT or Thyristor converter, and
IGCT or Thyristor inverter.

Applicable quadrants (Table 6.4) : 1 to 4. No energy saving as the drive is employed only for start-up of a system.

***Likely applications***
– Pumped storage schemes (like huge pumping of liquids to greater heights such as to a hill top)
– Gas turbine power plants
– Short-circuit generators for test benches
– Gas compressors
– Synchronous condensers
– Boiler feed pumps
– Propulsion drives

**(iv)** ***Cyclo converters***

In addition to the above inverter systems there is one more system in the family of frequency converters, called a cyclo-converter system. These drives are employed for very large, very low speed motors usually in MWs and use

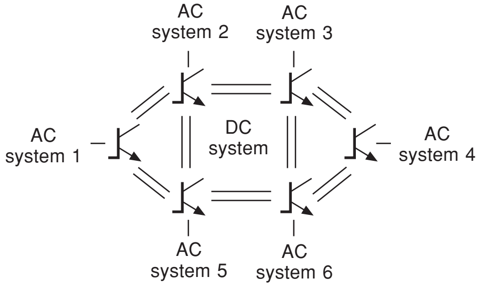



**Figure 6.22(k2)** Multi-terminal back to back ac-dc-ac configuration

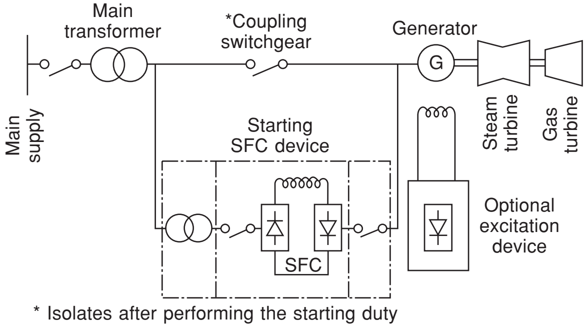
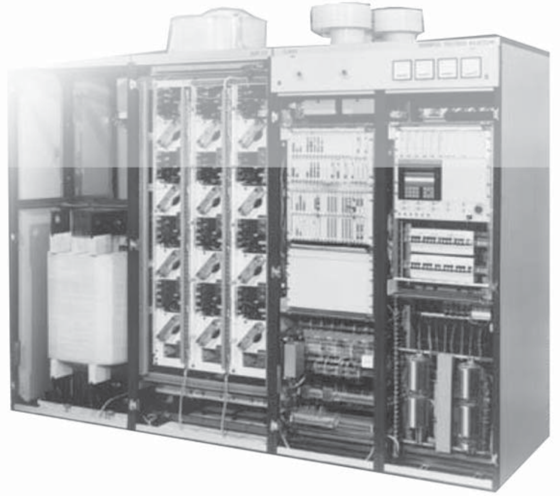



**Figure 6.22(l)** Scheme of a starting SFC for starting a gas turbine combined cycle (Courtesy: Alstom)

Front view with doors closed



Front view with doors removed

**Figure 6.22(m)** Starting SFC in a modular form (Courtesy: Alstom)

thyristor or GTO drives. IGBTs and IGCTs in such ratings may pose some limitation. It converts the fixed a.c. supply frequency to a variable frequency, lower than rated (0–100% f) directly and without rectifying it to a d.c. source. Better performance is achieved at lower speeds (0–(30–60) % f), while it is ideal at around 30% f. They are basically frequency converters. This system is more complex and expensive and has only limited applications, such as for very large squirrel cage, wound rotor or synchronous motors that are to operate at very low speeds (e.g. in cement plants and steel mills). For details see Further Reading.

### Circuit configuration (Figure 6.22(n)):

Thyristor or GTO frequency converter unit provides a variable frequency to the machine, for achieving speed variation 0–100% Nr.

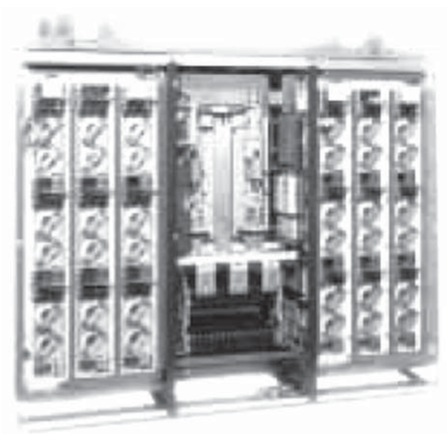


**Figure 6.22(n)** Circuit configuration of a thyristor based frequency converter unit (Cyclo converter) (Courtesy: Alstom)

Applicable quadrants (Table 6.4): 1 to 4 with active power management.

### 3. High current rectifiers

Electrolytic processes call for large and reliable supply of d.c. power. Current fluctuations in such processes adversely effect the efficiency and quality of the process. A short duration power loss may sometimes result in enormous loss of production. Likely applications of such rectifiers can be,

**(i)** ***Electrolysis processes:***
Electro-metallurgy – To refine aluminium, zinc, copper, cobalt, nickel and magnesium. Power requirement can be of the order of 20–500 kA, 100–1500 V d.c. (up to 750 MW).
Electro-chemistry – To produce chlorine, chlorate, hydrogen and graphite. Power requirement can be of the order of 15–250 kA, 100–400 V d.c. (up to 100 MW).

**(ii)** ***Arc furnaces and smelters***

*Circuit configuration* – like controlled rectifier units, Figure 6.24(a).
Figure 6.22(o) shows the view of a large rectifier unit with its cooling unit.
Devices used – Diodes, SCRs, GTOs and IGCTs or SGCTs

### 4. Static VAr compensators (SVCs)

These are employed for large fluctuating loads that may cause severe voltage fluctuations, line disturbances even outages. Such loads would demand for prompt reactive support to restore the normal supply conditions and save the system. Likely applications can be generating stations, power transmission and distribution networks,

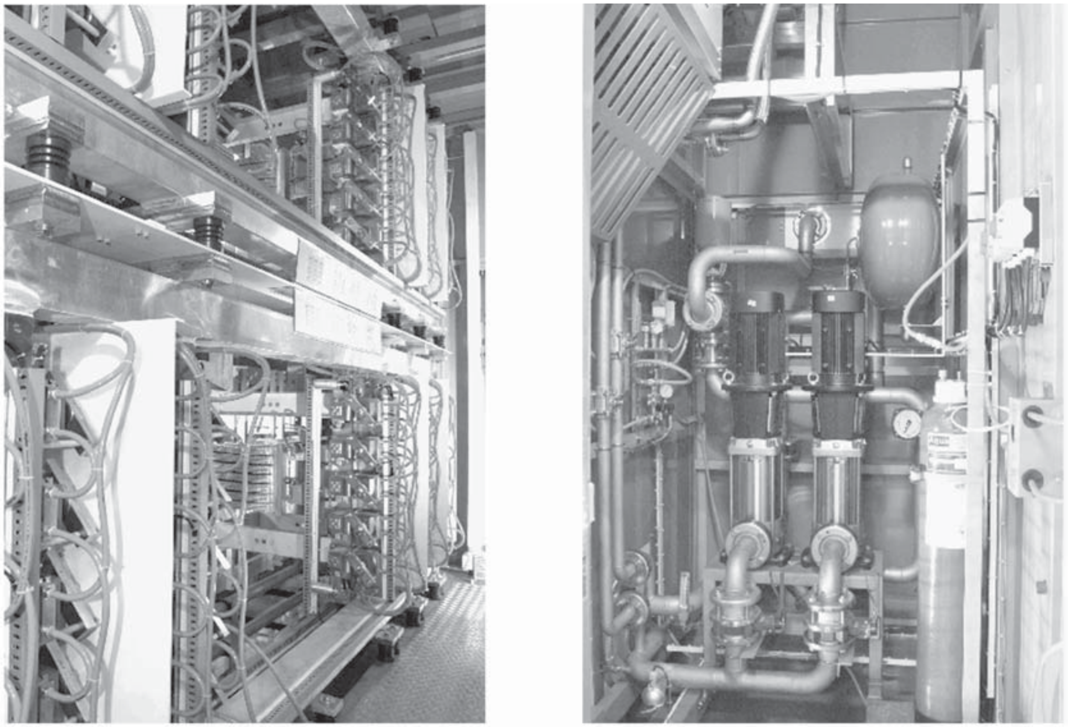



**Figure 6.22(o)** Rectifier unit 66kA / 1200 V at Pechiney Nederland (Courtesy: Alstom)

large loads like rolling mills, induction and arc furnaces for industrial heating and large rating rectifiers as noted above.

Figure 6.22(p1) shows a simple single line diagram of an SVC for a rolling mill. Part (i) of the figure illustrates the TCR portion and a filter circuit to absorb harmonic currents caused by TCR, while part (ii) shows the different harmonic filter circuits for the system and arc furnace harmonics. Figure 6.22(p2) shows the valve hall with thyristor power units for the TCR (thyristor controlled reactors). For details see Section 24.10. Figure 6.22(p3) shows the shunt reactors and Figure 6.22(p4) the capacitor banks for p.f. improvement and filter circuits installed in the open yard.

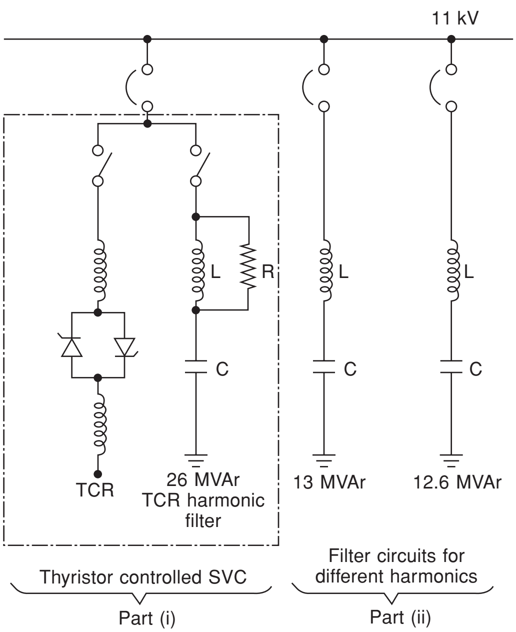



**Figure 6.22(p1)** Single line diagram of a typical SVC for a rolling mill (Courtesy: Alstom)

SVCs can also be employed for old substations that may have become over-utilized with time or are under-compensated and are ailing with perennial problems of flickering and outages. If the substations have space limitations or other constraints to accommodate SVCs at this stage, SVCs support can still be provided by mounting them on a mobile trolley placed at a convenient location to provide the needed reactive support promptly.

Device used – Mostly thyristors (SCRs).
For more details on SVCs see Section 24.10.

## 5. Static Drives

Depending on the rating and type of drive and its functional requirements one of the following combinations can be employed.

For converter unit – Power diodes, IGBTs, IGCTs, SGCTs, SCRs[a] or GTOs[a].
For inverter unit – IGBTs, IGCTs, SGCTs, SCRs[a] or GTOs[a].

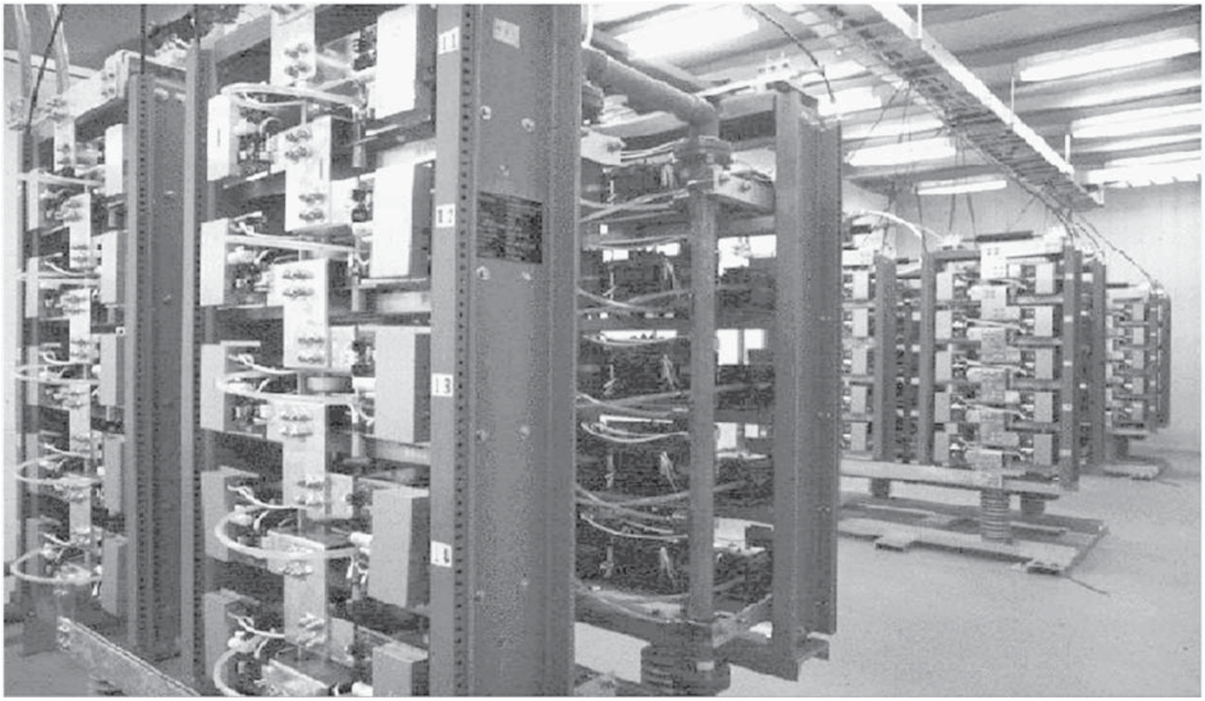


**Figure 6.22(p2)** View of a valve hall with thyristor power pack unit for the TCR (Courtesy: Alstom)

**Figure 6.22(p3)** SVC substation with shunt reactors in the foreground (Courtesy: Alstom)

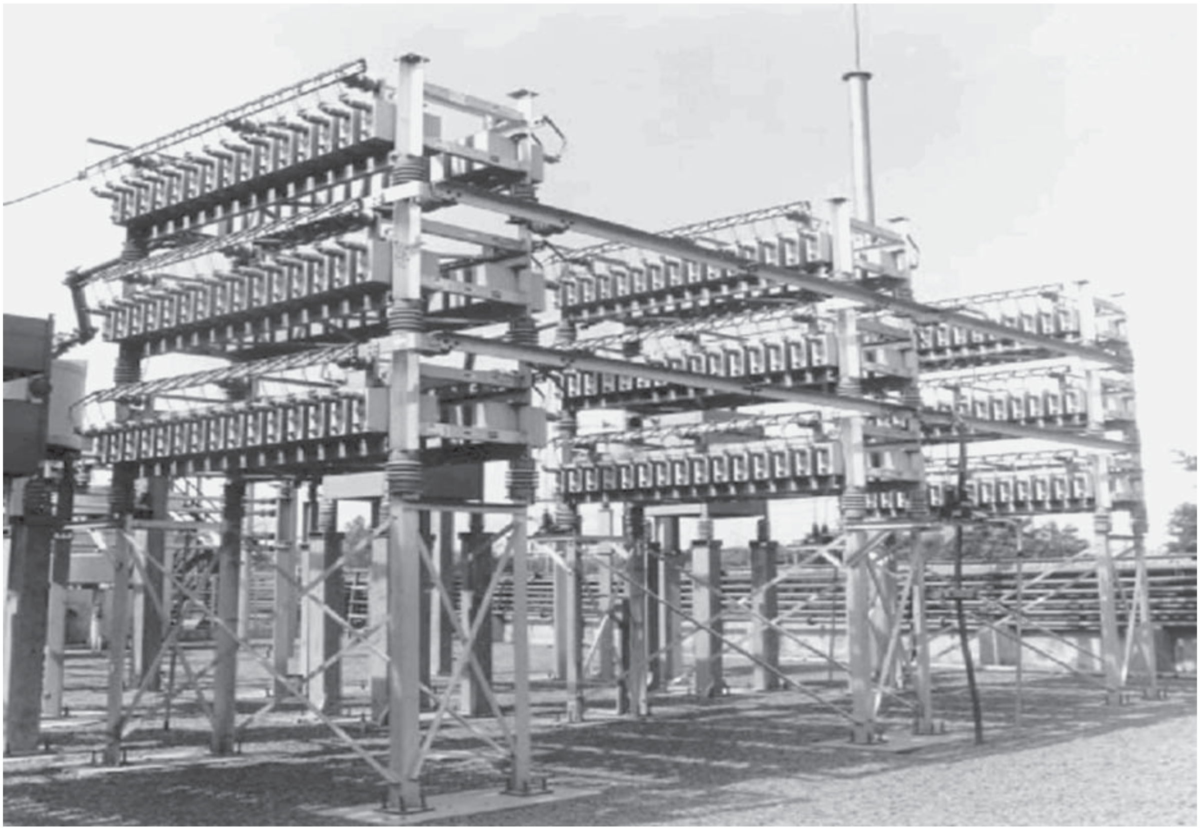


**Figure 6.22(p4)** PF control and SVC filter capacitor banks installed in the open yard (Courtesy: Alstom)

[a]*Note*
SCRs being line commutated are mostly used in converter circuits (for rectifiers) and GTOs being gate commutated in inverter circuits. Because of higher generation of harmonics their controls are not so accurate beyond a certain rating.

## Speed control

Use of *V/f* or FOC (phasor or vector control) and so also the use of open loop or closed loop control systems will depend upon the accuracy demand and the functional requirement of the motor these devices are controlling (for clarity see Table 6.2).

Drives up to a certain range (say up to 5 MW (although rating is no bar)) need not be custom built any more as the mathematical model of the drive can establish the motor parameters through auto-calibration and vary the same as desired using only a standard motor as discussed in Section 6.4.2. For accurate controls, however, only FOC must be used. Depending upon the field experiences and industrial requirements most manufacturers have developed their own standard drives that would suit most applications. Such drives are usually available off-the-shelf.

Custom built drives may become necessary only when it is known that the motor may have to operate under-loaded or is oversized and a smaller size of drive may serve the purpose (to cut on cost). Now the drive which is designed with parameters of a particular size of motor will not be able to auto-calibrate the data of a different size of motor. Such a situation may also arise when the drive is a retrofit to an existing conventional starter and the motor rating may not match the rating of the drive on similar grounds.

These drives are extensively used for control of a.c. and d.c. motors and synchronous motors. The usual devices used are noted in Table 6.2(b).

*Note*
The switching devices noted above are for a general reference only. A manufacturer may adapt to any of them depending upon his own development of newer devices and manufacturing practices.

## Compact sizes

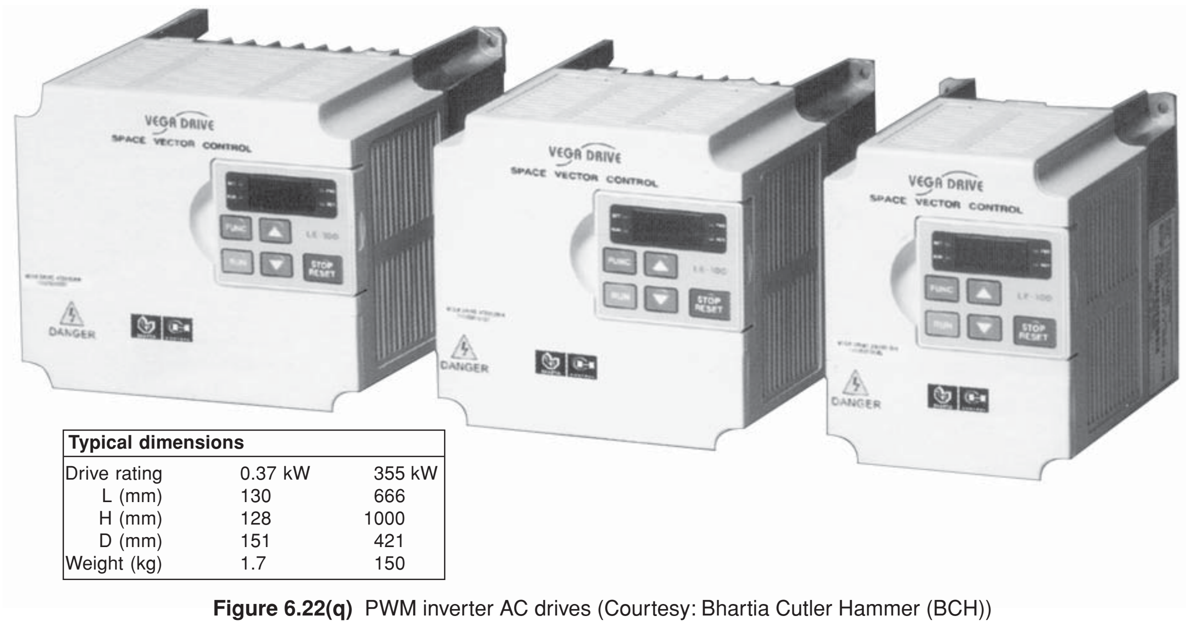
Such drives are extremely compact. Figure 6.22(q) illustrates a few drives in smaller ratings and their typical dimensions for a few ratings of a particular brand, for general reference.

## Remote drive and telemetry software

The drive is an intelligent electronic device (IED). It can be programmed and controlled locally via a PLC through a laptop computer (PC) and a keyboard as shown in Figure 6.22(r). It can also be made to communicate with a remote control station for

**Table 6.2b** Configuration of Static Drives

| *Parameters* | *Cage motors* | | | | *Wound motors* | | *Synchronous motors* | *D.C. motors* |
|---|---|---|---|---|---|---|---|---|
| | *When no active power management (APM) (feedback) is required* | | *When active power management (APM) (feedback) is required* | | *It would mean energy conservation* | | | *Discussed briefly in Section 6.19* |
| | *Rating up to a few MWs* | *Very large ratings* | *Rating up to a few MWs* | *Very large ratings* | *Rating up to a few MWs* | *Very large ratings* | *All ratings* | *All ratings* |
| (a) Converter unit – Devices Used | Power Diodes | Thyristors | IGBTs, IGCTs, SGCTs | GTOs, IGCTs, SGCTs | IGBTs, IGCTs, SGCTs | GTOs, IGCTs, SGCTs | Thyristors, IGCTs, SGCTs or GTOs | Anti parallel thyristors |
| (b) Inverter unit – Devices Used | IGBTs IGCTs | IGCTs, SGCTs, GTOs | IGBTs, IGCTs, SGCTs | IGCTs, SGCTs, GTOs | IGBTs, IGCTs, SGCTs | IGCTs, SGCTs, GTOs | Thyristors, IGCTs, SGCTs or GTOs | – |
| Speed control | With FOC (vector control) it is possible to vary the speed up to 1000 : 1 Nr with accuracy up to 25–100% Nr, ± 0.001% with closed loop control schemes (Table 6.2) | | | | | | – | As for cage motors |




| Typical dimensions | | |
|---|---|---|
| Drive rating | 0.37 kW | 355 kW |
| L (mm) | 130 | 666 |
| H (mm) | 128 | 1000 |
| D (mm) | 151 | 421 |
| Weight (kg) | 1.7 | 150 |

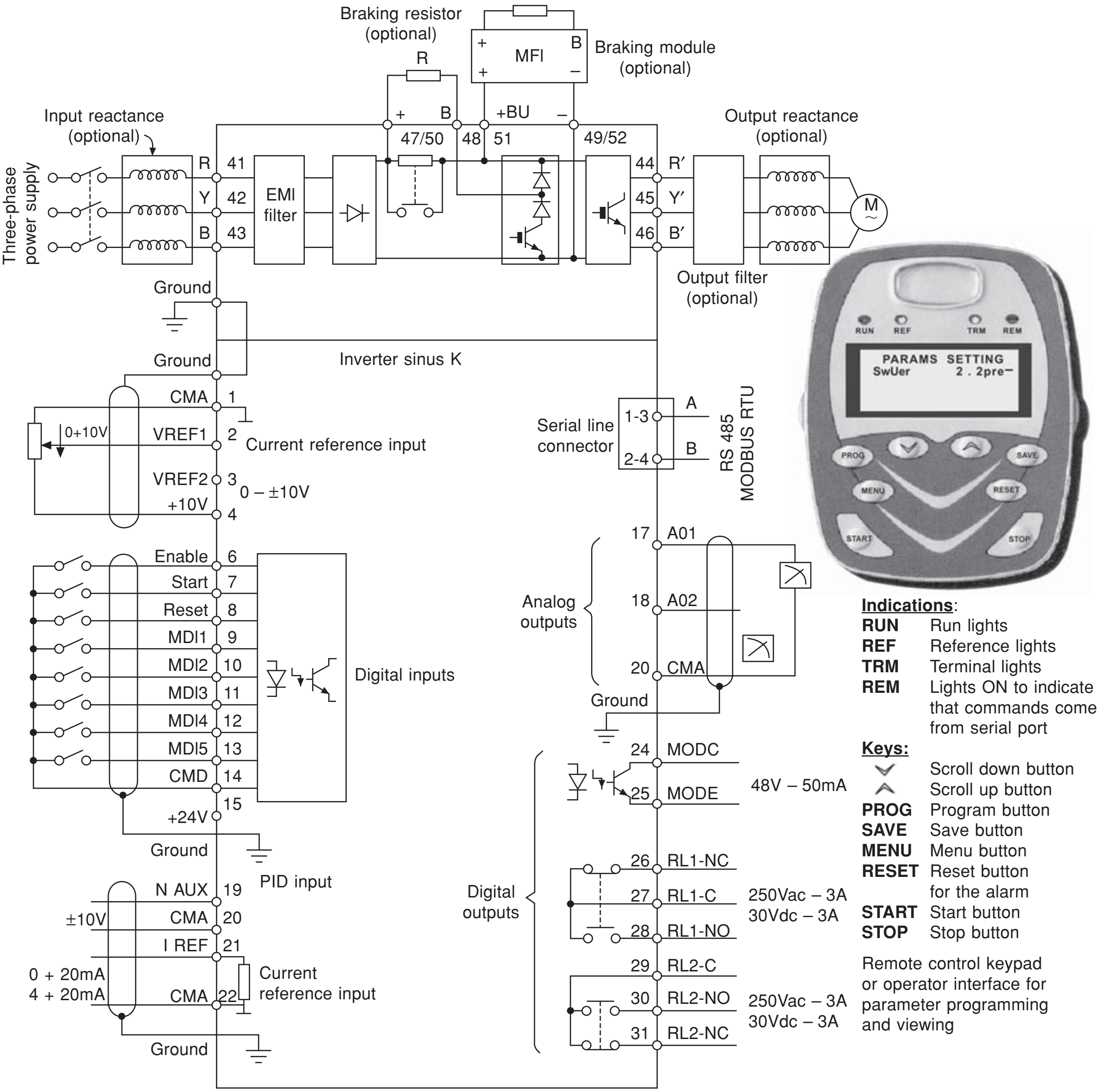
**Figure 6.22(q)** PWM inverter AC drives (Courtesy: Bhartia Cutler Hammer (BCH))

monitoring, control and automation, using communication protocols and media (field communication network) (Section 24.11.5). The local and remote control stations can set parameters, store, receive, transmit and save data from and on a PLC with keyboard emulation. It also allows remote service to the inverter via an internet connection through a modem. The remote drive software allows spontaneous communication between the user of the drive and the manufacturer's trouble-shooter desk or their technical support centre round the clock, to provide required assistance to the user in the hour of need.

Figure 6.22(r) gives a typical power and control terminals diagram of a PWM inverter a.c. drive along with terminal details and a remote control keypad for a general reference.

## 6.8 Conduction and commutation

A thyristor can be turned ON by the gate at any angle $\alpha$, with respect to the applied voltage waveform as shown in Figure 6.23(a) and (b) for half-wave and full-wave controlled rectifiers respectively. By varying the firing angle, which is possible through the firing circuit, the d.c. output voltage through a converter circuit can be varied, as illustrated in the figure. The voltage is full (maximum) when the firing angle is zero. Now the conduction angle is 180°. As the firing angle increases, the conduction angle decreases and so does the output voltage. The output voltage becomes zero when the firing angle becomes 180° and the conduction angle becomes zero. Thus the conduction, i.e. the power through a thyristor, can be varied linearly by varying the gate voltage




*Notes*

- Connection terminals of the braking resistor for different sizes of drives terminals 47 and 48, or 50 and 48
- Connection terminals of the external braking module for different sizes of drives terminals 51 and 52, or 51 and 49
- Terminals for inverter power supply from DC source: terminals 47 and 49

**Figure 6.22(r)** Typical power wiring diagram of a PWM inverter ac drive. The control terminal details are given below. (Courtesy: Bhartia Cutler Hammer)

Control terminal details

| *Terminal* | *Name* | *Description* | *I/O Features* |
|---|---|---|---|
| 1 | CMA | OV for main reference. | Control board zero volt |
| 2 | VREF1 | Input for voltage Vref1 main reference. | Vmax ± 10 V, Rin: 40 kΩ |
| 3 | VREF2 | Input for voltage Vref2 main reference. | Resolution: 10 bits |
| 4 | + 10 V | Supply for external potentiometer. | +10/Imax: 10 mA |
| 6 | ENABLE | Active input: inverter running with vector modulation (IFD) control for general applications Fluxed motor with sensorless vector (VTC) control for heavy torque performance Inactive input: in neutral regardless of the control mode. | Optoisolated digital input |
| 7 | START | Active input: inverter running. Inactive input: main ref. is reset and the motor stops with a deceleration ramp. | Optoisolated digital input |
| 8 | RESET | Active input: the inverter operation is reset after an emergency stop. | Optoisolated digital input |
| 9 | MDI1 | Multifunction digital input 1. | Optoisolated digital input |
| 10 | MDI2 | Multifunction digital input 2. | Optoisolated digital input |
| 11 | MDI3 | Multifunction digital input 3. | Optoisolated digital input |
| 12 | MDI4 | Multifunction digital input 4. | Optoisolated digital input |
| 13 | MDI5 | Multifunction digital input 5. | Optoisolated digital input<br>PTC with respect to BS4999 Pt. 111 (DIN 44081/DIN44082) |
| 14 | CMD | 0V optoisolated, multifunction digital inputs. | Optoisolated digital input input zero volt |
| 15 | +24 V | Auxiliary supply for optoisolated, multifunction digitial inputs | +24 V<br>Imax: 100 mA |
| 17 | AO1 | Multifunction analog output 1. | 0 ~ 10 V<br>Imax: 4 mA, 4–20 mA or 0–20 mA<br>Resolution: 8 bits |
| 18 | AO2 | Multifunction analog output 2 | 0 ~ 10 V<br>Imax: 4 mA, 4–20 mA or 0–20 mA<br>Resolution: 8 bits |
| 19 | INAUX | Auxiliary analog input. | Vmax: ± 10 V, Rin: 20 kΩ<br>Resolution: 10 bits |
| 20 | CMA | 0 V for auxiliary analog input. | Control board zero volt. |
| 21 | IREF | Input for current main reference (0–20 mA, 4–20 mA). | Rin: 100 Ω<br>Resolution: 10 bits |
| 22 | CMA | 0V for current main reference. | Control board zero volt |
| 24 | MDOC | Open collector digital output (collector terminal). | Open collector<br>NPN/PNP, Vmax: 48 V<br>Imax: 50 mA |
| 25 | MDOE | Open collector digital output (emitter terminal). | |
| 26 | RL1-NC | Multifunction digital relay output1 (NC contact). | 250 VAC, 3A<br>30 VDC, 3A |
| 27 | RL1-C | Multifunction digital relay output1 (common). | |
| 28 | RL1-NO | Multifunction digital relay output1 (NO contact). | |
| 29 | RL2-C | Multifunction digital relay output2 (common). | 250 VAC, 3A<br>30 VDC, 3A |
| 30 | RL2-NO | Multifunction digital relay output2 (NO contact). | |
| 31 | RL2-C | Multifunction digital relay output2 (NC contact). | |

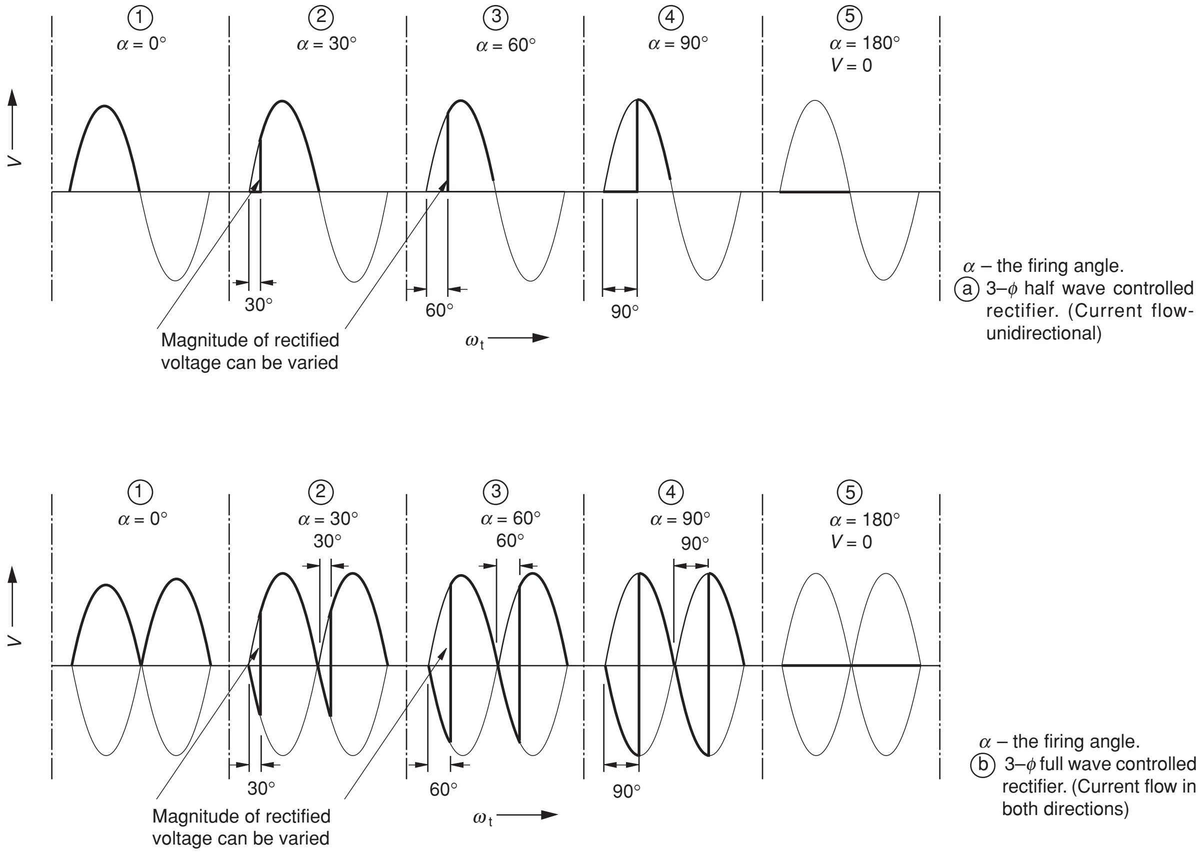



**Figure 6.23** Phase control of voltage through different gate firing angles

and its firing angle. Such control is termed phase control, and a rectifier or converter unit, employed to convert a.c. to a variable d.c., is called a controlled rectifier or a controlled converter.

In thyristor technology the switching OFF of a thyristor is conventionally termed commutation. In a.c. circuits, when the current through a thyristor passes through its natural zero, a reverse voltage appears automatically and turns OFF the thyristor. This is a natural commutation. No external circuit is now required to turn OFF the thyristor. They are therefore commonly called line commuting thyristors, like those used for a.c.–d.c. converters. But this is not so in d.c. circuits, as the current wave now does not pass through a natural zero. The forward current can now be forced to zero only through some external circuit to turn the thyristor OFF. This is termed forced commutation, such as when used for d.c.–a.c. inverters. Unlike a thyristor, a transistor can be switched OFF simply by removing the base signal and it requires no separate circuit to switch it OFF. In Table 6.3 we show a brief comparison between transistor devices and the basic thyristor (SCR). The triacs and GTOs and other thyristor devices fall in the same family, but with different features of conduction and commutation to suit different switching schemes. Other important characteristics are also shown in the table.

## 6.9 Circuit configurations of semiconductor devices

Semiconductor devices, as noted above, are used widely to convert a fixed a.c. supply to a fixed or variable d.c. supply, and then from a fixed d.c. supply to a variable a.c. supply, for example, for variable a.c. drives. A variable d.c. supply is used for the variable d.c. drives. The conventional nomenclature to identify the various types of these circuits and their applications is

- Converter or rectifier unit
- Inverter unit

### 6.9.1 Converter or rectifier unit

A converter unit is used for the control of d.c. machines and also to provide a d.c. source to an inverter unit controlling an a.c. machine. In d.c. drives the d.c. voltage after the converter unit should be variable, whereas for an a.c. drive it is kept fixed. The voltage is varied by the inverter unit. A converter unit is the basic power conversion scheme to convert an a.c. supply to a d.c. supply. Conventionally they are also known as rectifier units and can be arranged in four different modes to suit different applications of a motor as follows:

**Table 6.3** A brief comparison between a transistor and basic thyristor technology

| *Parameters* | *Basic transistor technology* | | | | | *Basic thyristor technology* | | | | |
|---|---|---|---|---|---|---|---|---|---|---|
| 1 As a static switch | Type of switch | Unidirectional | Bi-directional | Controllable | Uncontrollable | Type of switch | Unidirectional | Bi-directional | Controllable | Uncontrollable |
| | Power BJT and power Darlington | ✓ | – | ✓ | – | SCR | ✓ | – | ✓ | – |
| | Power MOSFET | – | ✓ | Controllable in forward and uncontrollable in reverse direction | | GTO | ✓ | – | ✓ | – |
| | | | | | | TRIAC | – | ✓ | Controllable in both directions | – |
| | IGBT | ✓ | – | ✓ | – | Reverse Conducting thyristor | – | ✓ | Controllable in both directions | – |

*Note:* The above switching devices by themselves or in conjunction with power diodes have been developed into a variety of new devices to suit any power conversion and control application. MCTs and IGCTs are a few such hybrid devices. See Section 6.7.3 and Table 6.2a.

| *Parameters* | *Basic transistor technology* | *Basic thyristor technology* |
|---|---|---|
| 2 Switch OFF characteristics | This calls for no switching OFF circuitry, since a transistor can be switched OFF simply by removing the base signal. | Once fired, a thyristor cannot be controlled. It requires a forced commutation to switch it off and the gate control is quite cumbersome. To switch OFF, the conducting current is reduced to less than its holding current. The commutation circuitry is therefore highly complex and also influences the reliability of a thyristor application. |
| 3 Controls | These require only the base signal to switch ON. Thus providing a simpler technology. | There are six thyristor firing circuits required for a 3Φ system, as there are two thyristors connected back to back each phase. The whole scheme is therefore complex and less reliable. |
| 4 For conduction in both directions | Generally two circuits are used. | These can be connected in anti-parallel. |
| 5 Switching frequency | Power MOSFETs (now being used as a control circuit switching device) and IGBTs can handle much higher switching frequencies, compared to a thyristor. In an a.c. motor control, fast switching is mandatory (when using PWM technique in the inverter circuit) and therefore transistors are preferred. | Very low switching frequency, but a GTO is suitable for frequent switchings. |
| 6 Rating | (a) Can handle only moderate currents and voltages. A BJT is used mostly in electronic control circuits. As they are small, hundreds can be placed on a small PCB (printed circuit board).<br>(b) IGBTs alone are used for power applications. See Section 6.7.2. | IGBTs and their hybrids can handle much larger powers. Typical $V$ and $f$ ratings for each unit can be upto and even higher than $V \simeq 6$ kV[a] and $I \simeq 6$ kA[a] (see Table 6.2a) |
| 7 To vary $V$ and $f$ | Both $V$ and $f$ can be varied with the help of pulse width modulation (PWM) in the inverter circuit. The converter unit normally is an uncontrolled power diode rectifier. | By varying the gate firing angle, $V$ can also be varied. With SCRs frequency variation is not possible. *Note:* Since an inverter is not line commutated, the SCRs have a switching limitation and hence a limitation in frequency variation. When thyristors are to perform frequent switchings, GTOs are used in the inverter circuit. |
| 8 Heating effect | Compared to thyristors high heat dissipation. | With continuous developments and improvements in design and construction, their heat loss has been greatly contained. But their relatively low switching frequency capability makes IGBTs, IGCTs and SEGTs as more preferred choices. |
| 9 Cost factor | Much more economical compared to a thyristor drive in this range. | In smaller ratings they are economical. In an a.c. to d.c. converter, for instance, for the control of a d.c. motor, where a variable d.c. voltage is desirable, SCRs are used extensively and the voltage variation is obtained by varying the firing angle. Since the SCRs are now line commutated, they pose no switching OFF problem. |

[a] Subject to applicable deratings. See also Table 6.2a. These ratings can be enhanced by connecting them in series-parallel combinations. In series to enhance the voltage rating and in parallel to enhance the current rating. But the controls may not be so accurate as with a single device.

**1 Uncontrolled rectifier units** These can be

- Half wave similar to Figure 6.24(a) configurations (B) and (C), using one diode per phase instead of a thyristor, or
- Full wave similar to Figure 6.24(a) configuration (A), using two diodes in anti-parallel per phase instead of thyristors or in the form of a centre-point configuration.

Figure 6.24(b) shows a thyristor controller unit.

**2 Controlled rectifier units** These can also be

- Half wave (Figure 6.24(a)), configurations (B) and (C) using one thyristor per phase, or
- Full wave (Figure 6.24(a)), configuration (A), using two thyristors in anti-parallel per phase or in the form of a centre-point configuration.

*Note*
A half-wave rectifier is a single-bridge rectifier and is suitable for only single-quadrant operations I or III. A full-wave rectifier is a double-bridge rectifier and suitable for multi-quadrant operations particularly quadrants II and IV. See Table 6.4.

The uncontrolled rectifier units are simple rectifiers and use power diodes universally. The rectification obtained is uncontrollable and provides only a fixed d.c. voltage output from a fixed a.c. supply. The diodes have no control over their switching instants. These rectifier units are used extensively to provide a fixed d.c. source to an inverter unit when being employed to control an a.c. machine. Since there is no switching of diodes involved, there are no voltage surges across the diodes. There is thus no need to provide a snubber circuit across the diodes to protect them against such surges, as discussed later.

When, however, a variable d.c. voltage output is required, controlled rectifiers are used. Now the diodes are replaced by one or two phase-controlled thyristors (SCRs) in each phase, one thyristor per phase for a half wave and two per phase in anti-parallel for a full wave rectification. The required voltage variation is achieved by adjusting the delay time of the gate-firing pulse ($\alpha$) to each thyristor unit. Thyristor circuits are possible up to any rating by arranging them in series-parallel combinations. With the advances in this technology, it is possible to achieve a total matching of switching sequences of all the thyristors to ensure an accurate and fully cohesive operation (switching of all thyristors at the same instant). Unlike diode, a thyristor does not turn ON automatically at the instant it becomes forward biased and requires a gate pulse at its gate terminal to switch it ON. This is obtained through a control circuit which is a part of the rectifier unit. The gate-firing pulse is provided at the appropriate instant to each thyristor in each switching cycle, positive half to negative half, to obtain the required voltage. The switching is automatic as the thyristors are line commutated and at each half cycle the voltage waveform passes through its natural zero. The instants may be delayed from 0° to 180° electrical to obtain the required infinite control in the output supply, as illustrated in Figure 6.23(b). The d.c. output is controlled by adjusting the delay time of the gate-firing pulse. A few voltage waveforms are illustrated in Figure 6.23(b) at different firing angles. Phase control of positive and negative half waves of each phase can be infinitely varied to meet any power demand.

A half-wave rectifier is able to provide only a uni-directional d.c. power source which may also contain many a.c. ripples (Figure 6.24(a)). A full-wave rectifier is employed to reduce such ripples, on the one hand, and provide a d.c. power in forward as well as reverse directions, on the other. A fixed forward and reverse d.c. power is required for an inverter unit when it is employed to control an a.c. machine. Now an uncontrolled rectifier unit is adequate as $V$ and $f$ control is obtained through the inverter unit.

A controlled rectifier unit is necessary when it has to control a d.c. machine, which would call for a variable d.c. voltage. When the d.c. machine has to operate in only one direction (quadrants I or III) a half-wave controlled rectifier will be adequate and when the machine has to operate in either direction, a full-wave controlled rectifier will be essential.

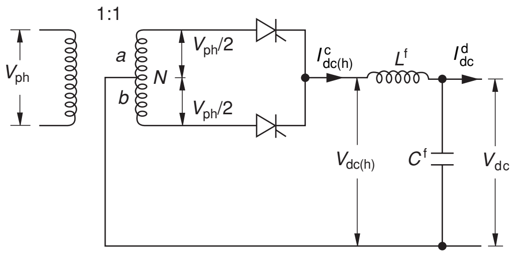
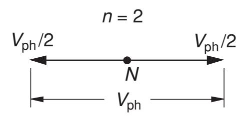
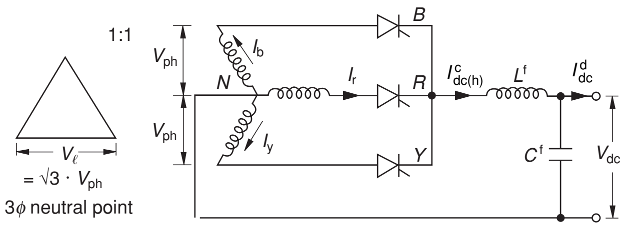
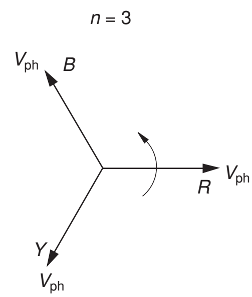
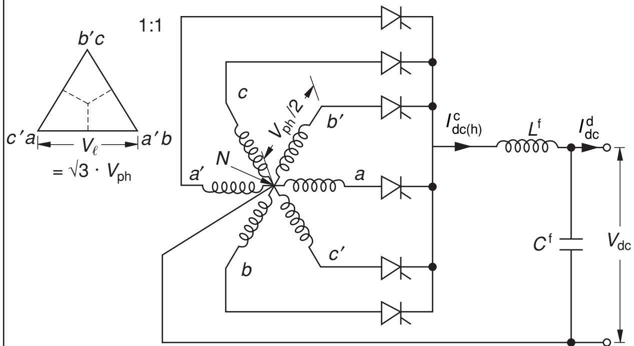
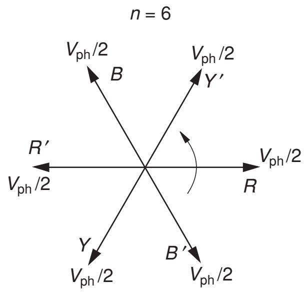
For operations in quadrants II and IV it is essential to have an unrestricted flow of reverse power and hence an additional feedback inverter unit would also be essential, as shown in Figures 6.31–6.33, depending upon the configuration of the converter unit.

*Note*
It is possible that at some locations there is no a.c. source available, such as for battery-operated lifts and motor vehicles. Such applications may also call for a variable d.c. source. When it is so, it can be achieved with the use of a chopper circuit which uses the conventional semiconductor devices. The devices are switched at high repetitive frequencies to obtain the required variation in the output voltage as with the use of a phase-controlled thyristor rectifier. A typical chopper circuit is shown in Figure 6.25, using diodes and a controlled uni-directional semiconductor switch, which can be a thyristor or an IGBT.

### *Relevance of different quadrants of a converter unit*

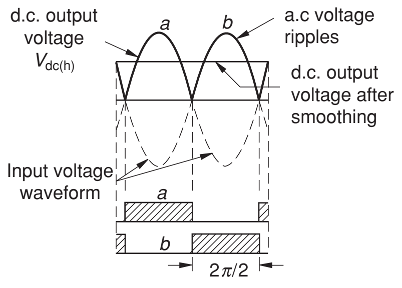
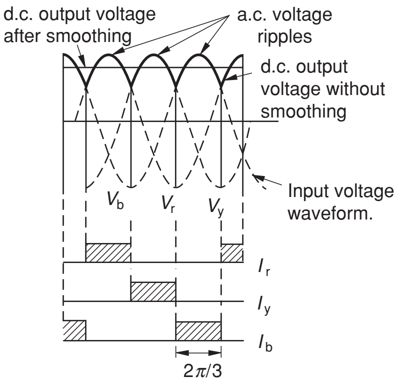
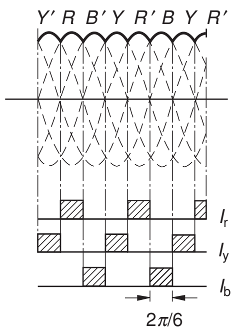
Depending upon the mode of operation of a motor, the type of converter unit can be decided. For simplicity, the operation (conduction) of a motor can be represented by four quadrants as illustrated in Table 6.4.

- **Quadrant I** Both $V$ and $T$ are positive. The machine can be run only in one direction (say, forward). Braking operations are possible. It is a converter mode and either a half-wave or a full-wave rectifier can be used.
- **Quadrant II** Now $T$ is in the reverse direction and the machine can be run in the reverse direction. Braking and regeneration are possible. For regeneration an additional bridge will be essential as discussed later. For current to flow in either direction, a full-wave rectifier will also be essential.
- **Quadrant III** Now both the voltage and the torque are in the reverse direction otherwise it is similar to Quadrant I. The machine can now be run in the reverse direction.
- **Quadrant IV** Now the voltage alone is in the reverse direction and the machine can be run in the forward direction. Braking and regeneration are possible. The machine in the forward direction can generate, when it is brought down from a higher speed to a lower

| *Thyristor configuration* | *Winding arrangement and thyristor [a] rectifier configuration* | *No. of phases (pulse number, n and voltage phasor)* |
|---|---|---|
| (A) Bi-phase (Single phase centre point) (Full wave rectifier) |   |   |
| (B) 3-phase, (Half wave rectifier) |   |   |
| (C) 6[e]-phase (Half wave rectifier) |   |   |

*Notes* a Considering the firing angle $\alpha = 0$, the values are same for power diodes also.
b (i) This ratio is true for power diodes. Or thyristors with $\alpha = 0$,
(ii) $V_{ac}$ is considered as the phase voltage of each transformer limb. Even $V_{ph}/2$ is considered as $V_{ac}$, for simplicity.
(iii) To derive this equation refer to Vithayathil (1995).
c $I_{dc(h)}$ – d.c. output with ripples

**Figure 6.24(a)** *(Contd.)*

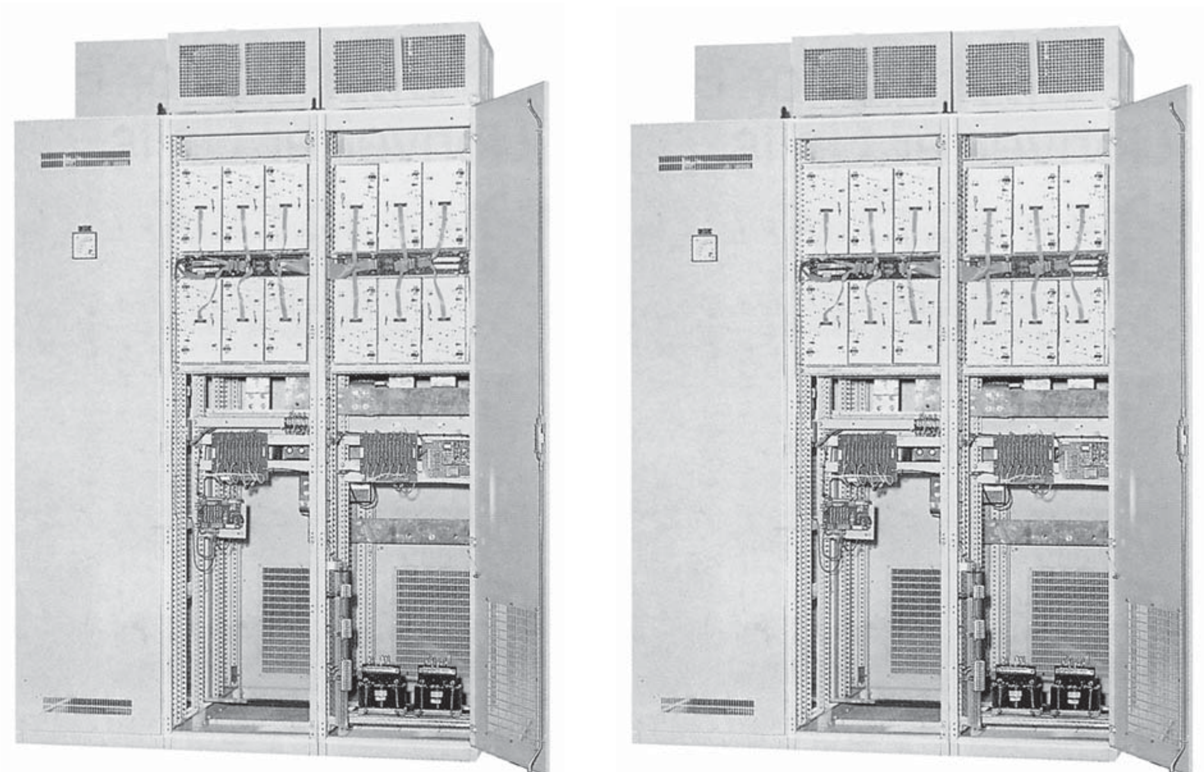
| | Voltage and current waveforms for $\alpha = 0$ | Output voltage with firing angle $\alpha$ $V_{dc}^{b} = \frac{\sqrt{2}\cdot n}{\pi} \times V_{ac}\sin\frac{\pi}{n}\cos\alpha$ | Ratio of d.c. to a.c. voltage $\frac{V_{dc}^{b}}{V_{ac}} = \frac{\sqrt{2}\cdot n}{\pi}\sin\frac{\pi}{n}\cos\alpha$ for $\alpha = 0$ | Current on the a.c. side (tryristor[a] current) | | |
|---|---|---|---|---|---|---|
| | | | | $I_{av} = \frac{I_{dc}}{n}$ | $I_{rms} = \frac{I_{dc}}{\sqrt{n}}$ | $I_{peak} = I_{dc}$ |
| (A) | <br>Current output waveform. Each phase conducting for 2π/2 = 180° | $\frac{2\sqrt{2}V_{ac}}{\pi}\cos\alpha$ | 0.9 | $\frac{I_{dc}}{2}$ | $\frac{I_{dc}}{\sqrt{2}}$ | $I_{dc}$ |
| (B) | <br>Current output waveform each phase conducting for 2π/3 = 120° | $\frac{3\sqrt{6}V_{ac}}{2\pi}\cos\alpha$ | 1.17 | $\frac{I_{dc}}{3}$ | $\frac{I_{dc}}{\sqrt{3}}$ | $I_{dc}$ |
| (C) | <br>Current output waveform. Each phase conducting for 2π/6 = 60° | $\frac{3\sqrt{2}V_{ac}}{\pi}\cos\alpha$ | 1.35 | $\frac{I_{dc}}{6}$ | $\frac{I_{dc}}{\sqrt{6}}$ | $I_{dc}$ |

d $I_{dc}$ – d.c. output with minimal ripples
e The six secondary phases are obtained by shorting the centre points of each of the three-phase windings of a 3$\phi$ transformer secondary.
f For use of $L$ and $C$ refer to Figure 6.34.

**Figure 6.24(a)** A few configurations of controlled rectifier units (for uncontrolled rectifier units the thyristors (SCRs) are replaced with diodes)

**Table 6.4** Operation of a motor in different modes and the corresponding conducting quadrants of a controlled converter unit

| Reverse-running quadrants (−T) | Forward-running quadrants (+T) |
|---|---|
| **Quadrant II** (+V)<br>Operations – driving and braking (reverse motoring) – reverse regeneration<br>(N, T)<br>Configuration – only a full wave – an additional bridge for regeneration<br>Mode of operation – when motoring – converter when regenerating – inverter | **Quadrant I** (+V)<br>Operations – driving and braking (forward motoring)<br>Configuration – half wave or full wave<br>Mode of operation – converter<br>(N, T) |
| **Quadrant III** (−V)<br>Operations – driving and braking (reverse motoring)<br>Configuration – half wave or full wave<br>Mode of operation – converter<br>(T, N) | **Quadrant IV** (−V)<br>Operations – driving and braking (forward motoring) – forward regeneration<br>Configuration – only a full wave – an additional bridge for regeneration<br>Mode of operation – when motoring – converter – when regeneration – inverter<br>(T, N) |



Rear view

Front view

**Figure 6.24(b)** Thyristor cubicles for high-power static converter units (Courtesy: Siemens)

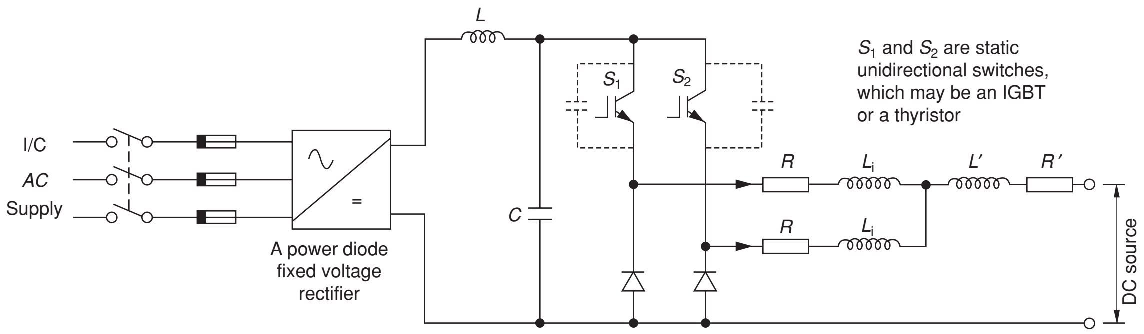



**Figure 6.25** A typical bi-phase chopper scheme

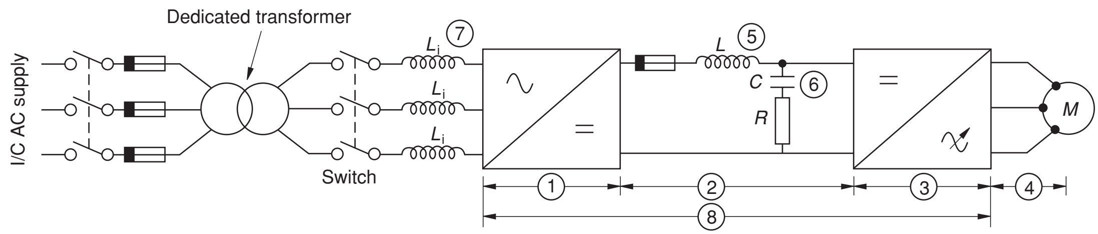
speed or when feeding a load going down hill etc. For regeneration, an additional bridge will be essential as noted above, and for the current to flow in either direction, a full-wave rectifier.

The changeover from motoring to regeneration, i.e. from Quadrant I to Quadrant IV or from Quadrant III to Quadrant II or vice versa, is achieved by first bringing the torque to zero by ceasing the firing of one bridge and commencing that of the other.

### 6.9.2 Inverter unit

The purpose of an inverter unit is to invert a fixed d.c. power to a variable a.c. power which can be achieved in two ways:

1 *Using IGBT devices* These are the latest in the field of static power control. They are easy to handle and control, besides being inexpensive in the presently developed ratings compared to GTOs. A fixed d.c. voltage is obtained through a power diode converter. The conversion from a fixed a.c. supply to a variable a.c. supply is thus cheap and easy to handle. A complete system composed of a converter and an inverter unit, as used to control an a.c. motor, is more commonly called an inverter. In Figure 6.26(a) we show a basic IGBT or thyristor (GTO) inverter unit. Single IGBTs developed so far can handle machines up to 1.4 MVA. There are some deratings of an IGBT on account of a lower r.m.s. value of the inverted power, compared to the power input to the inverter, the configuration of the inverter (which defines the inverted waveform), the ambient temperature, safety margins etc.




(1) *Rectifier or converter unit*
Converts a.c to d.c. and can be a power diode fixed voltage for a.c. drives (phase controlled thyristor converter for d.c. drives)
(2) D.C.link circuit connecting (1) to (3)
(3) *Inverter unit*
Inverts d.c. to variable a.c. and can be an IGBT, or a thyristor (GTO) circuit.
(4) Variable *V* with an VSI (Figure 6.28(a)) or variable I with a CSI (Figure 6.29), and variable *f* a.c. power output
(5) *Inductor L*
 (i) It is necessary to smooth a.c. ripples, whether it is a power diode or a thyristor rectifier. It also suppresses harmonics when, (1) is a controlled rectifier, producing a.c. ripples and harmonics
 (ii) Can be a large size inductor, when (3) is a current source inverter (CSI) (Figure 6.29)
 (iii) Provides a short-circuit protection for a fault in d.c. link, by adding to its impedance
(6) Charging capacitor, to hold the charge, by smoothing the output ripples and providing a near constant voltage source to the inverter circuit, when it is a voltage source inverter. When the converter is a thyristor converter, a resistance *R* is also provided with *C* to make it suitable to perform its duty under frequent thyristor switchings, by quickly discharging it through *R*. Now it becomes a snubber circuit, to also protect the inverter devices from $dv/dt$.
(7) (a) Current limiting reactor on I/C side, to control $di/dt$ during switching of thyristor units when (1) is a phase controlled rectifier. Its rating is usually 2–4% of the source impedance (Section 6.13.1). Not necessary when the unit is supplied through a dedicated transformer.
 (b) Also required to limit $di/dt$ to the solid-state circuits when the source of supply is large and is protected by current limiting device (Section 6.13.2, Figure 6.35).
(8) Inverter unit (conventional name). Converts fixed a.c. to variable a.c.

**Figure 6.26(a)** Basic IGBT or thyristor (GTO) inverter unit

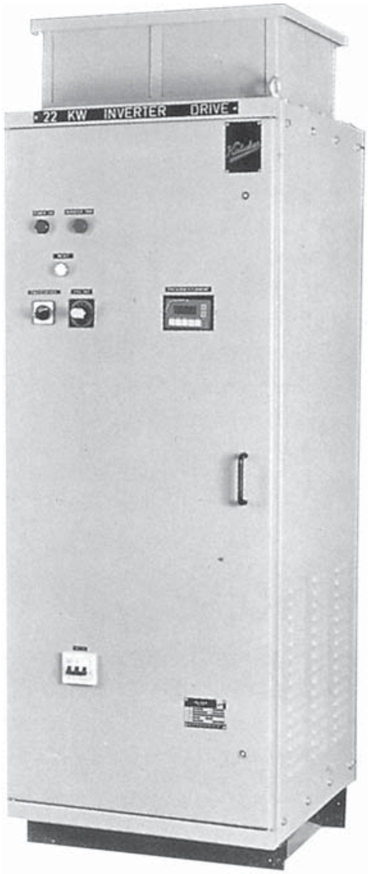



**Figure 6.26(b)** A small rating IGBT inverter unit (Courtesy: Kirloskar Electric)

Their over-current capacity is defined by the over-load current and its duration. IEC 60146-1-1 has defined it as 150–300%, depending upon the application, for a duration of one minute. The manufacturer can derate a device for a required load cycle (over-loading and its duration) according to the application. Unless specified, the present normal practice is to produce such devices for an over-current of 150% for one minute. A higher starting current than this or a start longer than one minute may call for a higher derating. Figure 6.26(b) shows a small rating inverter unit.

IGBTs can be used for still higher ratings as shown in Table 6.2a by connecting them in series-parallel combination. For still higher ratings, thyristors (GTOs) are preferred for better reliability. More number of IGBT in series-parallel combination may sometimes act erratically and perform inconsistently. The latest in the static power devices is the hybrids of transistor and thyristor families as noted already.

2 *Using thyristor devices (GTOs)* Thyristor (GTO) inverter circuits are used for very large machines and handling large powers such as reactive power management, HVDC transmission and large rectifiers etc.

### *To obtain variable V and f*

*In IGBTs through pulse width modulation (PWM)* The frequency in the inverter circuit is varied by frequently switching the IGBTs ON and OFF in each half cycle. While the voltage is controlled with the help of pulse width modulation, which is a technique for varying the duty cycle or the zeros of the inverter output voltage pulses. The duty cycle or CDF (cyclic duration factor) of the pulses is the ratio of the period of actual conduction in one half cycle to the total period of one half cycle.

For Figure 6.27(a)

$$\text{CDF} = \frac{t_1 + t_2 + t_3 + t_4 + t_5 + t_6}{T} \tag{6.3}$$

where $t_1$, $t_2$, … $t_6$ are the pulse widths in one half cycle. If $V$ is the amplitude of the output voltage pulses, then the r.m.s. value of the output a.c. voltage

$$(V_{\text{r.m.s.}})^2 = (\text{CDF}) \cdot V^2$$

$$\text{or } V_{\text{r.m.s.}} = V \cdot \sqrt{\text{CDF}} \tag{6.4}$$

By varying the CDF, i.e. the pulse widths of the a.c. output voltage waveform, the output, $V_{\text{r.m.s.}}$ can be varied.

The CDF can be controlled by controlling the period of conduction, in other words, the pulse widths (periodic time period, $T$ remaining the same). Thus the a.c. output

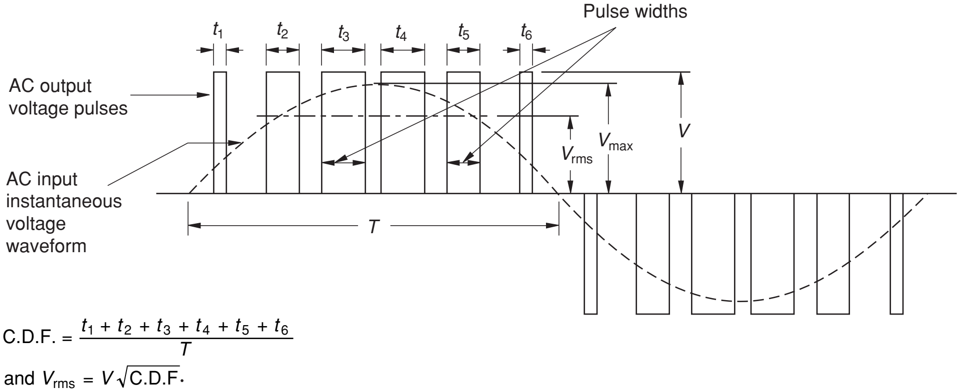



**Figure 6.27(a)** Varying the output a.c. voltage with PWM technique

voltage in an IGBT inverter can be controlled with the help of modulation. The modulation in the inverter circuit is achieved by superposing a carrier voltage waveform of much higher frequency on the natural voltage waveform. Figure 6.27(b) is a simple block diagram for a PWM scheme, the natural voltage being the voltage obtained by the switching of the IGBTs. The carrier wave can be of any shape, the frequency of which is altered, to obtain the required degree of modulation, and hence the voltage, while the amplitude is kept fixed. The amplitude is a matter of scheme design. (For more detail refer to the textbooks in the Further Reading.) Generally, a triangular wave is used as shown in Figure 6.27(b) to obtain a more uniform sinusoidal voltage waveform. By Fourier analysis we can establish the amplitude of voltage and quality of waveform (distortions), and by controlling the pulse widths through the frequency of the carrier wave, we can decide the best modulation to obtain the required amplitude and a near sinusoidal output voltage waveform. (For details of Fourier analysis, refer to a textbook.) This is the most commonly used technique in the inverter circuit to obtain the required *V/f* pattern. It is also economical and can be used to control multi-motor drives through a single unit. To obtain an accurate *V/f* control, it is essential that the voltage is maintained uniform (without ripples) as much as possible. This can be achieved by providing a capacitor across the d.c. link as shown in Figure 6.26(a). The purpose of the capacitor is to hold the charge and smooth the output a.c. ripples of each diode and hence provide a near-uniform d.c. voltage. The charge retained by a capacitor can be expressed by

$$Q = C\,\frac{dv}{dt} \tag{6.5}$$

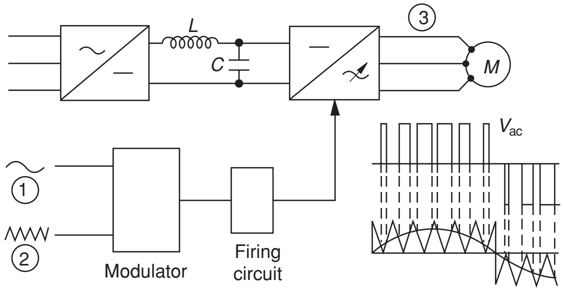
where

$Q$ = charge stored by the capacitor unit

$C$ = capacitance of the capacitor

$\frac{dv}{dt}$ = rate of voltage change or a.c. ripples in the d.c. link

The higher the value of $C$, the lower will be the voltage overshoots in the rectified voltage and the inverter circuit




① Reference voltage
Inverter natural voltage waveform before modulation, improved to a near sinusoidal waveform, with the use of *L* and *C*.
② Carrier voltage
Triangular voltage waveform of fixed amplitude
③ Variable frequency and modulated voltage output (*V/f*) as desired.

**Figure 6.27(b)** Varying the output a.c. voltage with PWM technique

would be fed by an almost constant voltage source. The capacitor in the circuit also provides an indirect protection from the voltage surges.

The above method is used to vary the frequency and the voltage of the inverter output (motor side) according to the process needs, irrespective of the electronics scheme adopted to obtain the required speed control.

*Note*
The variation of frequency is generally up to its fundamental value, i.e. 0–50 or 0–60 Hz, in view of the fact that the motor is generally required to operate below the base speed. At higher frequencies the motor will overspeed, for which its own suitability as well as the suitability of the mechanical system must be pre-checked. When, however, such a situation be desirable (field weakening region, Figure 6.7(b)), the frequency may be varied to the desired level by switching, keeping the output voltage to the rated value. Since the torque of the motor will now reduce $\alpha 1/N$, this must be checked with the load requirement.

### *In GTOs*

The frequency in the inverter circuit is varied by switching the GTO pairs ON and OFF repeatedly through their gate control in each one half cycle. The rate of frequency variation will depend upon the frequency of switching of the GTO pairs. The voltage variation is obtained by varying the gate firing angle, $\alpha$.

By using converter–inverter combinations in different configurations and by applying a proper gate control, a variety of fixed and variable output parameters of a fixed parameter a.c. input power can be obtained. When the motor is operating at very low speeds, say, below 5% of $N_r$, the motor voltage demand is also low. If the inverter circuit is load commutated (motor side), its phase current will have to commutate with a very low voltage at the load side. It is difficult to guarantee reliable commutation at such low voltages. Pulse width commutation is therefore also employed in thyristor drives when the motor has to operate at very low speeds. Where the motors are very large, cyclo converters can also be employed. Below we discuss a few inverter configurations. Generally PWM (for IGBTs) and gate control schemes (for thyristors) may be applied to these inverter circuits to obtain the required variable a.c. supply parameters at the output line to suit a particular requirement.

## 6.9.3 Voltage source inverter (VSI)

This is the most commonly used inverter for the control of a.c. motors and is shown in Figure 6.28(a). The fixed d.c. voltage from the uncontrolled rectifier converter acts as a voltage source to the inverter. Since the variation is based on voltage, the inverter is called a voltage source inverter (VSI). The voltage in the inverter unit is varied to the required level by using a pulse width modulation, as noted earlier. Through the switching circuit of the inverter the frequency of the output supply is varied by repeated switching of the IGBTs. The frequency of switching of the IGBTs determines the frequency of the output a.c. voltage and is a matter of system design and practices adopted by a manufacturer. For a general reference it may be in the range of,

– Low rating motors say, up to 200 kW – 8–16 kHz

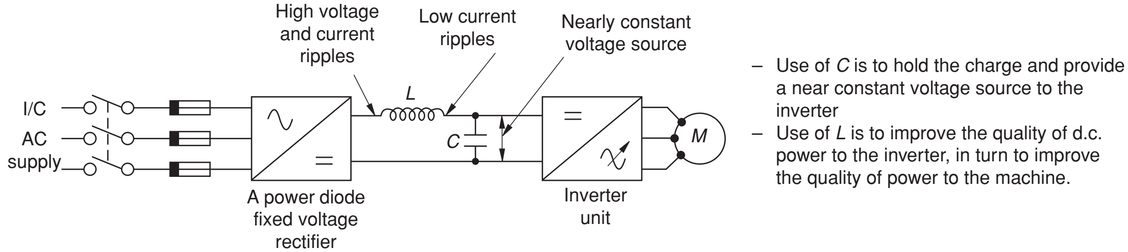



**Figure 6.28(a)** A voltage source inverter (VSI)

– Medium rating motors say, up to 1 MW – 4–8 kHz
– Large motors above 1 MW or so – 2–4 kHz

The inverter unit (converter–inverter unit combined) can be considered as comprising

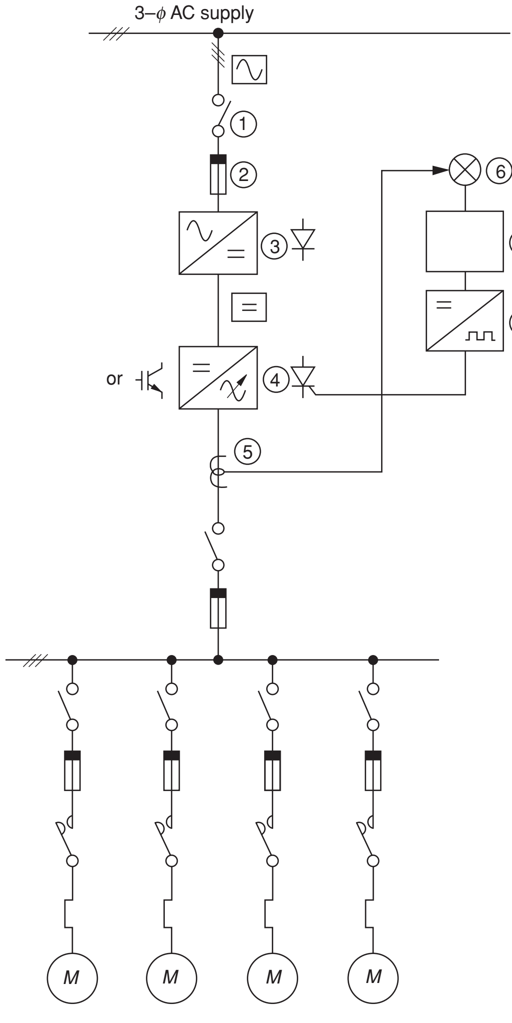
1 **Rectifier unit (converter)** This is a fixed voltage uncontrolled diode bridge rectifier.
2 **Smoothing circuit** To obtain a near-constant voltage source for the inverter circuit a smoothing capacitance across the d.c. link is used to smooth the $\frac{dv}{dt}$ ripples present in the d.c. link after conversion. The capacitor retains the charge and provides a near-constant d.c. voltage output.
3 **Inverter unit** This inverts the fixed d.c. voltage to a variable $V$ and $f$ a.c. voltage. Such a system can control multi-motor drives, operating on the same bus and requiring similar speed controls as illustrated in Figure 6.28(b). The inverter parameters can be closely controlled with the help of feedback controls and sensing devices.

### 6.9.4 Current source inverter (CSI) (to vary $I_\ell$ and $f$)

This is similar to a voltage source inverter, except that now it is the rectified current that is varied rather than the voltage. On the input side of the inverter it acts as a current source. Since this current is already pre-set for the required a.c. output current $I_\ell$, the motor current is always within its permissible loading limits. The current control, therefore, provides self-control to the motor. As before, as shown in Figure 6.29, a fixed d.c. voltage is provided through an uncontrolled diode bridge rectifier. This voltage is converted to a constant current source with the help of a large series inductor in the d.c. link. The purpose of this inductor is to provide a near-constant current source by reducing the d$i$/d$t$ ripples. The inductor plays the role of a current source and acts like a current chopper. Since, the voltage across the inductor can be expressed by

$$V \simeq L\frac{di}{dt} \text{ (ignoring } R \text{ of the circuit)} \qquad (6.6)$$

the higher the value of $L$, the lower will be the current overshoots, i.e. the rate of the current change d$i$/d$t$, through the inductor. A high value of the inductor would make it possible to provide a near-constant current source for the inverter circuit. With more modifications, the above can be made to operate according to a pre-defined current waveform for very accurate speed control of a motor.




① Isolator
② Fuse
③ Diode bridge rectifier (converter)
④ Inverter unit IGBT or thyristor depending upon the size of machine
⑤ CT
⑥ Current comparator
⑦ Current amplifier and controller
⑧ Gate control in case of thyristor inverters only.

**Figure 6.28(b)** Single line block diagram showing cascade connections of motors on a variable voltage common bus, using a VSI

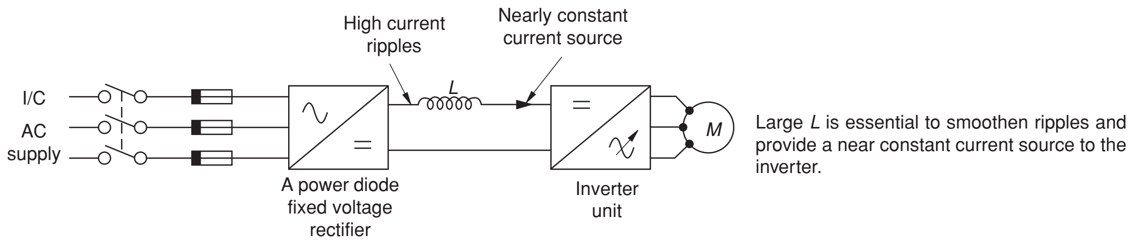



**Figure 6.29** A current source inverter (CSI)

Now they may be called current regulated inverters. With feedback controls, precise control of a motor can be achieved. A current source inverter provides a simpler and better control and may be preferred for large drives, particularly where regenerative controls are involved.

Now the frequency of the a.c. output current is also varied through the switching of the IGBTs in the inverter unit, as noted earlier, and the current is varied by varying the output a.c. voltage, using the same PWM technique as for a VSI. Through this scheme only single-motor control is possible, as different motors will have different currents, as they may be of different ratings. However, it is more suitable for larger drives, as it is easy to handle currents rather than voltages. Only limitation is poor dynamic response.

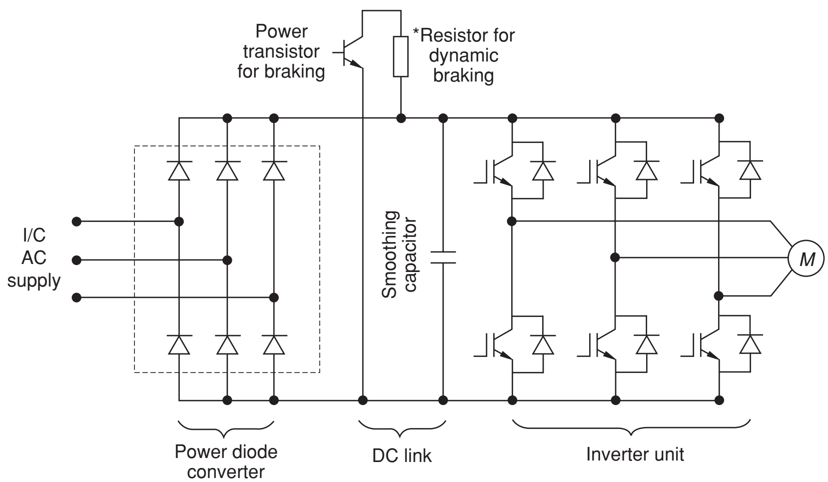
### 6.9.5 The regenerative schemes

A motor can fall in a generator mode when the machine is energized and is run beyond its synchronous speed, such as when driving a load, travelling downhill or when its speed is reduced to perform a specific duty. The same conditions will appear when a running machine is reversed, whether it is an a.c. or a d.c. machine.

In any of the above generating modes if the surplus energy is not fed back to the supply source it may have to be dissipated in some other form. Otherwise it may raise the d.c. link voltage beyond its acceptable level, and lead to an unwanted trip of the machine or overheating. It may also endanger the static devices used in the inverter circuit or the components in the d.c. link. One simple way to do this is to consume these in a resistor as shown in Figure 6.30. This is known as dynamic braking, and the regenerative energy is wasted. The resistor is introduced in the circuit through a bus voltage sensor. As soon as the bus voltage rises beyond a pre-set limit, the resistor is switched into the circuit. In smaller motors it is common practice to dissipate the heat of regeneration in this way but in larger machines it can be a substantial drain on the useful energy, particularly when the machine is called upon to perform frequent variations of speed, reversals or brakings. It is, therefore, advisable to conserve this energy by feeding it to the other drives or by transferring it back to the source of supply, which can be done in the following ways.

When controlling an a.c. machine, the converter is usually a full-wave, power diode fixed-type rectifier and the *V/f* is controlled through the IGBT inverter. For regenerative mode, the d.c. bus is connected in anti-parallel with a full-wave voltage-controlled inverter as shown in Figure 6.31.

During a regenerative braking, the d.c. voltage starts to rise, which the inverter regulates to the required level (*V* and *f*) and feeds back the regenerative energy to the source of supply. This is known as synchronous inversion. When the inverter is of a lower voltage rating than the




* Mechanical braking or non-regenerative braking:
For small brake power, resistance unit is small and can be located within the main enclosure. But for higher power that may call for large resistance units and have to dissipate excessive heat, it is mounted as a separate unit. The resistance units are short-time rated depending upon the duty they have to perform.

**Figure 6.30** An IGBT inverter unit with dynamic braking

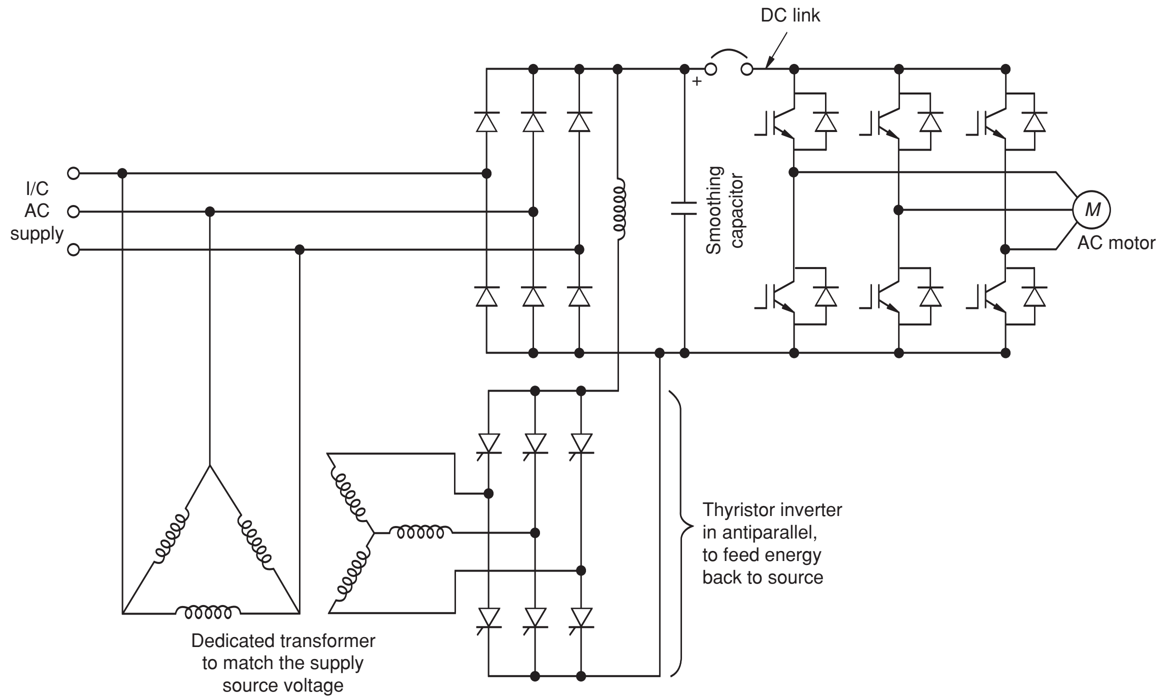



**Figure 6.31** Regenerative energy feedback arrangement for an inverter unit

supply source, a transformer between this and the supply source as shown will also be necessary to regulate the feedback voltage to the required level. The delta side of the transformer may be connected to the supply side to eliminate the third harmonic quantities to the source. Over-voltage, over-load and short-circuit protections may

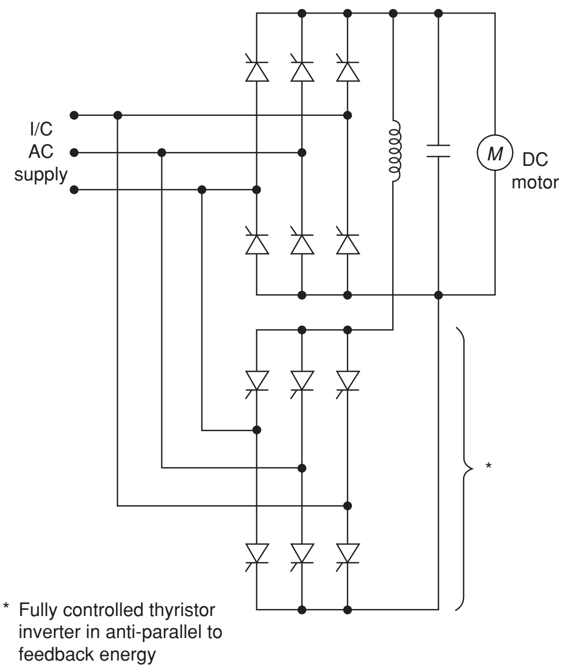



**Figure 6.32** Regenerative energy feedback arrangement for a converter unit

be provided on the d.c. bus, on which may occur a fault or whose voltage may rise beyond the pre-set limits. A similar inverter circuit is used in a d.c. machine also when the regenerative energy is to be fed back to the source of supply (Figure 6.32).

It is also possible to use an IGBT converter instead of a power diode or a thyristor converter. A single IGBT converter unit is capable of performing both the jobs, converting the fixed a.c. supply to a fixed or variable d.c. voltage to control an a.c. machine and during a regenerative mode, feeding the regenerative energy back to the source of supply. A separate transformer will not be necessary now, as the same IGBT circuit will act as a regenerative inverter. The switching of an IGBT is now an easy feature. The harmonics are also too low, and the p.f. can be maintained up to unity. Now the IGBT converter can be called a sinusoidal converter, as it will provide a near sinusoidal waveform of voltage and currents during a feedback. The d.c. bus can also be made a common bus to feed a number of drives through their individual IGBT inverters to cut on cost. This is possible when a number of such drives are operating in the vicinity or on the same process line. Figure 6.33 illustrates a simple scheme with a common d.c. bus.

## 6.10 Smoothing ripples in the d.c. link

A power diode rectifier unit feeding a fixed d.c. power to an inverter unit to control an a.c. motor. Or a thyristor rectifier unit, directly controlling a d.c. motor. Both contain a.c. ripples in their d.c. outputs, as illustrated in Figure

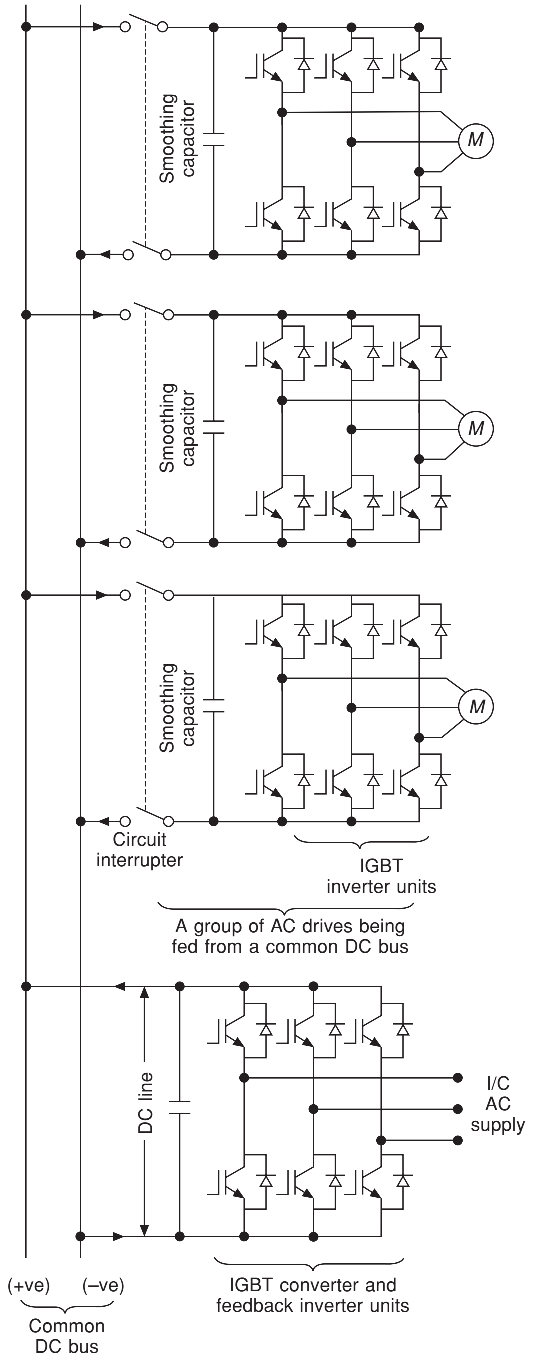



**Figure 6.33** An IGBT converter-cum-inverter unit to feed back regenerative energy

6.24(a). It is essential to smooth these ripples to improve the quality of d.c. power. To achieve this, a series inductor *L* is provided in the d.c. link as shown in Figures 6.24(a) and 6.28(a). In the process it also reduces the harmonics on the input side. To cut on cost, it is possible to limit the use of such inductors to say, 10 h.p. and above. In smaller drives the ripples may not significantly influence the performance of the machine. Nevertheless it is advisable to suppress the harmonics to retain a healthy supply side of the system.

The inductance in the d.c. link may cause a reverse voltage spike across the power diodes or thyristors as a result of the decay of the reverse current (release of its stored energy). A power device may be protected against such voltage spikes through an R–C snubber circuit, as shown in Figure 6.37. (Snubber circuit is discussed later.)

## 6.11 Providing a constant d.c. voltage source

After smoothing the d.c. voltage may contain moderate ripples ($dv/dt$) not desirable when a constant voltage d.c. source is needed. To achieve this, a charging capacitor *C* is also provided across the d.c. link for all sizes of drives as shown in Figures 6.24(a) and 6.28(a).

## 6.12 Providing a constant current source

Instead of a charging capacitor *C*, a large size series inductor *L* is introduced in the d.c. link (Figure 6.29). Since $V = L\, di/dt$, the larger the value of *L*, the lower will be the current overshoots ($di/dt$) and a near-constant d.c. link current source is obtained for the inverter unit.

## 6.13 Generation of harmonics, over-voltages and voltage surges in a static device switching circuit and to mitigate their effects

### 6.13.1 Harmonics generated by phase-controlled rectifier units

A switched static device (particularly a thyristor) produces voltage and current transients similar to inductive or capacitive switching (Section 17.7). They also produce harmonics. However, a power diode converter unit having no switching sequence is devoid of such a phenomenon. A thyristor (SCR) switched phase-controlled converter unit produces large quantities of harmonics on the supply (a.c.) side, also in the d.c. link and also voltage and current surges on the incoming supply side. All these factors are not desirable and must be suppressed or tamed right at the point of occurrence to save the connected equipment and the devices. Keeping in line with EMC/EMI regulations (Section 23.18) it is mandatory on the part of all those who generate electro-magnetic (EM) disturbances to suppress the same to the desired level. IEEE-519 and G5/4 (UK Engineering recommendations) provide the recommended values for each harmonic (see Tables 23.2 and 23.3). Similarly the electronic circuits must also be protected and made compatible with EMI effects. See IEC 61800-3 for static drives. Below we discuss these phenomena and their possible remedies.

### *Current harmonics on the incoming a.c. supply side*

- The presence of harmonic quantities in the electronic circuit distorts the sinusoidal incoming supply system to a non-sinusoidal one, the magnitude of which will depend upon the configuration of the converter circuit and the variation in the connected load. The line side converter unit draws a somewhat squarish waveform current from the mains, as analysed in Figure 23.7. It may adversely influence the power equipment operating on the incoming supply side of the system, which may be a motor, a transformer or a generator, due to higher no-load losses as a consequence of high harmonic frequencies (Equations (1.12) and (1.13)) and higher level of electromagnetic noise. It may also cause over-loading of the capacitor banks connected on the incoming side and subject all these equipment to higher voltage stresses. The higher inductive loading also diminishes the p.f. of the system. To contain the influence of these features, the use of filter circuits to suppress the harmonics and power capacitors, to improve the system p.f. on the incoming side are mandatory to maintain a healthy supply system, particularly when it is feeding a few phase-controlled converter units, handling large machines and generating high harmonics. Figure 6.34 shows the use of an inductor in the incoming circuit to suppress the harmonics and limit current overshoots. Power capacitors are not shown, which can be provided for the whole system at a centralized location. The design of filter circuits and the size of power capacitors, to adopt a more appropriate corrective method, will require a meticulous network analysis to determine the actual numbers and magnitudes of such harmonics present in the system. The subject is dealt with in more detail in Section 23.5.2. In Figure 6.24(a) we have shown a few more common types of thyristor configurations, their voltage and current waveforms and the application of reactors to suppress the harmonics and smooth a.c. ripples. As a rule of thumb in absence of network analysis the inductor is usually considered about 2–4% of the source impedance (it is preferred to keep it as low as possible to avoid extra inductive loss and voltage drop). This inductor is not necessary when the unit is supplied through a dedicated transformer.
- Phase-controlled rectifier circuits generate excessive odd harmonics such as 5th, 7th, 11th, 13th etc., depending upon the pulse number, $n$ of the circuit and configuration adopted, as discussed in Section 23.6(b). These harmonics add to the inductive loading of the circuit since $X_L \propto f_h$ and diminish the p.f. of the system, although they hardly influence an induction motor, (Section 23.5.2(B)). The 3rd harmonics are totally absent because mostly six pulse thyristor converters are employed, which eliminate all the 3rd harmonics from the voltage and the current output waveforms. Thyristors in other configurations such as 12, 18 and 24 pulses are also possible, which can eliminate most of the harmonics from the input and output waveforms. This is also known as phase multiplication method using one transformer and two secondaries for 12 pulse or three secondaries for 18 pulse or four secondaries for 24 pulse rectifier units. Figure 6.34(a) shows a 12 and 18 pulse converter circuits respectively. The higher the pulse number, the closer it approaches the mean and effective (r.m.s.) values of the rectified voltage and the voltage approaches a near peak value (Section 23.6(b)). (See also Figure 23.10.) However, higher pulse thyristor converters become very expensive and are employed only for very large power applications.

### *Current harmonics in a d.c. link*

To limit the current harmonics generated in the d.c. link, series smoothing reactors are inserted on the d.c. side as shown in Figure 6.24(a). They are large iron core unsaturable reactors (L). (For details on reactors see Chapter 27.) They provide high impedance paths to the different harmonic quantities and suppress the more prominent of these at the source, and provide a near smooth d.c. output voltage waveform. For large power applications, requiring a near-constant d.c. output, more accurate L–C circuits (even more than one) may be provided in the d.c. link to suppress the more prominent of the harmonic quantities.

A large inductor in the d.c. link may also play the following roles:

1 In the event of a fault in the d.c. link it will add to the circuit impedance and limit the rate of rise of fault current, since under a transient condition

$$V = L\frac{\mathrm{d}i}{\mathrm{d}t}$$

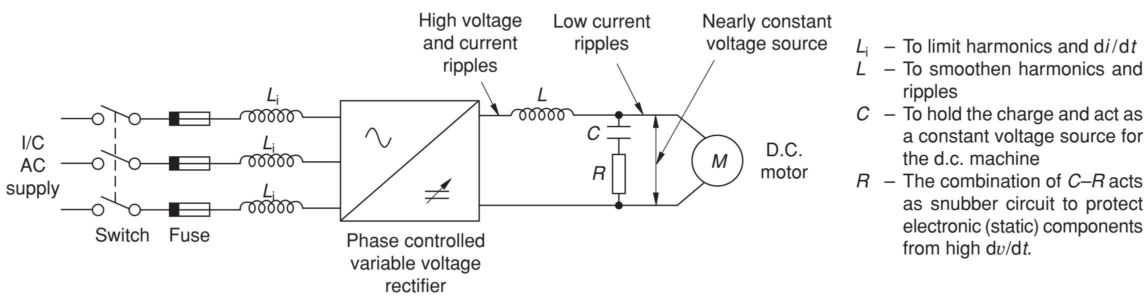
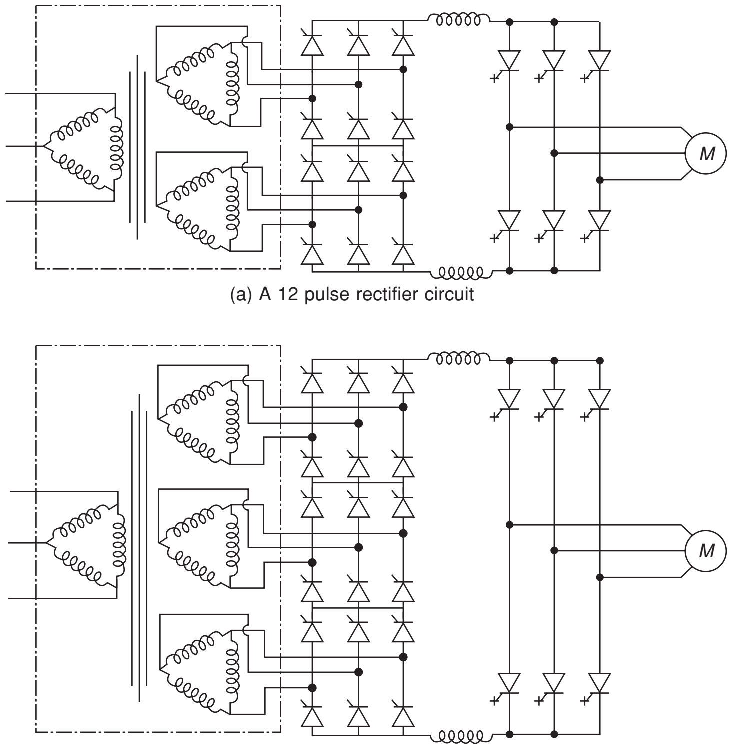



**Figure 6.34** Application of inductor and capacitor with a controlled bridge rectifier (for control of d.c. machines)




(a) A 12 pulse rectifier circuit

(b) An 18 pulse rectifier circuit

**Figure 6.34(a)** High pulse rectifier circuits to reduce harmonics on the I/C supply side (Courtesy: Allen Bradley)

A high value of $L$ will limit the rate of rise of fault current for the same voltage and save the circuit components.

2 A low d$i$/d$t$ will also absorb the current over-shoots and help smooth the d.c. link current waveform.

3 It has a disadvantage as it may have a sluggish response to the control circuit demands due to its high time constant ($\tau = L/R$)

## 6.13.2 Over-voltages and voltage surges

A voltage transient generally is a phenomenon of HV systems, as discussed in Chapter 17. Moderate long-duration switching surges (voltage spikes), other than lightning and the transference of surges however, are noticed on an LV system also as discussed in Section 17.7.6. See also IEEE-C62.41. Some of it may be due to switchings of static devices and affect motors and semiconductor devices connected on it.

- LV systems that are prone to frequent faults and outages or constitute a number of inductive or capacitive switched loads and welding transformers may generate temporary over-voltages (TOVs) and voltage surges. Such systems must be studied carefully, and when felt necessary, a metal oxide varistor (MOV) or a large inductor be installed at the incoming side of the semiconductor circuits.
- Fault condition particularly when the LV distribution is fed through a large transformer and the outgoing feeders are protected by current limiting devices, HRC fuses or breakers. In the event of a fault, on a large outgoing feeder, as illustrated in Figure 6.35, the protective device will trip and trap an electromagnetic

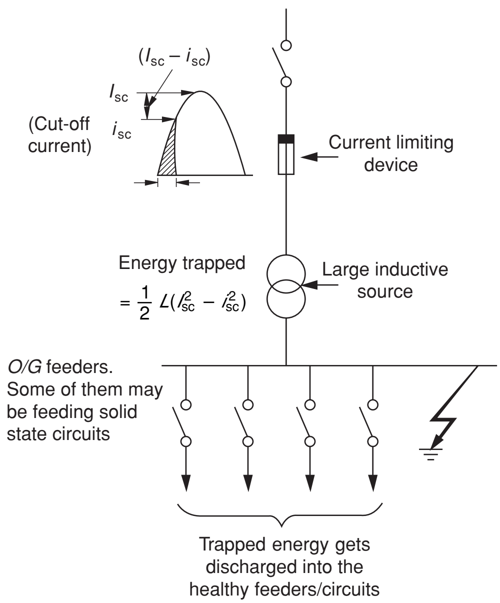



**Figure 6.35** Trapped energy distribution of a large feeding source during a fault clearing by a current-limiting device

(EM) charge within the transformer and the inter-connecting cables, with an energy equivalent to $\frac{1}{2} L \cdot (I_{sc}^2 - i_{sc}^2)$. $L$ is the inductance of the transformer and the inter-connecting cables up to the static circuits and $i_{sc}$ the cut-off current of the prospective fault current $I_{sc}$, at the instant of fault. The clearing of fault by a current limiting device is a transient condition of the transformer and is synonymous with a switching condition and may generate switching surges. This energy is discharged into the circuits located down stream of the feeding source. The trapped energy is a source of danger to all healthy circuits that are located near the source, and the worst affected are the feeders, that may be switched at such instants.

All the above surges and even the transference of a lightning surge from an overhead line through inter-connecting cables are uni-directional and reflect in nearly full at a junction and cause a doubling effect, hence they are more dangerous. The semiconductor devices can be saved from such harmful effects by absorbing the trapped energy. The effect of a trapped charge is somewhat a replica of a discharge of a surge arrester. In a surge arrester, the energy above the protective level of the arrester is discharged through the ground (Section 18.5). In this case, it is discharged into the healthy circuits down stream. Normal practice to tackle such a situation is to provide an inductor $L_2$, sufficient in size, to absorb this energy at the receiving end of the static circuit as illustrated in Figure 6.36. This protection is applicable to all types of electronic circuits. It is equally applicable even in a power diode converter unit, involving no switching operations.

### *Voltage surges within the converter unit*

This is applicable to thyristor (SCR) circuits to protect all the semiconductor devices used in the switching circuit, such as diodes (also power diodes) or IGBTs, in addition to SCRs. The same protection can be applied to all the semiconductor circuits likely to experience high d$v$/d$t$.

The role of SCRs is to vary the supply parameters, which require frequent changes in $V$, i.e. d$v$/d$t$ and in $I$, i.e. d$i$/d$t$ in an energized condition. Because of momentary phase-to-phase short-circuit, d$v$/d$t$ occurs during switching OFF and d$i$/d$t$ during switching ON sequences. Both are transient conditions and may damage the semiconductor devices used in the circuit. To protect the devices, the transient conditions can be dealt with as follows:

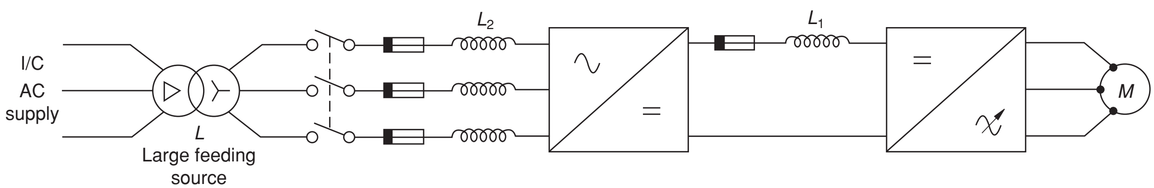
- ***Voltage transients* (d*v*/d*t*)** When a thyristor switches from a closed to an open condition, i.e. from a conducting to a non-conducting mode, a transient recovery voltage (TRV) appears. This is a transient condition and the rate of change of voltage can be expressed by

$$Q = C' \cdot \frac{dv}{dt}$$

where
$Q$ = charge stored within the devices before occurrence of the switching
$C'$ = leakage capacitance of the thyristor between its junctions
d$v$/d$t$ = rate of rise of recovery voltage (r.r.r.v.)

During a switching OFF sequence this charge must be dissipated quickly otherwise it may cause dangerous over-voltages, which may damage the static devices used in the circuit or turn them ON when this is not wanted. A thyristor is switched ON only when there is a current pulse applied to its gate. It is possible that the gate may turn ON without this pulse as a result of excessive forward d$v$/d$t$ due to leakage capacitance between the thyristor junctions. The leakage capacitance may cause a charging current through the gate. When it exceeds its threshold value, it can turn the gate ON. d$v$/d$t$ is therefore a very important limiting parameter to avoid an erratic turn ON of the thyristors, which may corrupt the output parameters and lead to malfunctioning of the whole system or cause a short-circuit and damage the static devices used in the circuit. It is therefore important to suppress such transients within safe limits. It is possible to contain them, provided that the stored energy can be dissipated quickly into another source. This is achieved by providing a snubber circuit across each static device as noted below, similar to the use of a quenching medium in an HV interrupter (Section 19.2).

- ***Snubber circuit*** More conventional protection from high d$v$/d$t$ is to provide an $R - C$ circuit across each device, as shown in Figure 6.37. The circuit provides a low impedance path to all the harmonic quantities and draws large charging currents and absorbs the energy released, $Q$, and in turn damps d$v$/d$t$ within safe limits across each device. Now $Q = C$ (d$v$/d$t$) where $C$ is the capacitance used across the device. During a turn OFF operation the stored energy, $Q$, of




$L$ – Inductance of the supply source
$L_1$ – Inductor to smooth ripples
$L_2$ – Inductor to absorb the trapped energy up to $\frac{1}{2} L \cdot (I_{sc}^2 - i_{sc}^2)$ [partly absorbed by the feeder's own impedance and other feeders connected on the same line]

**Figure 6.36** Use of inductor on the supply side of a static drive to absorb the trapped energy

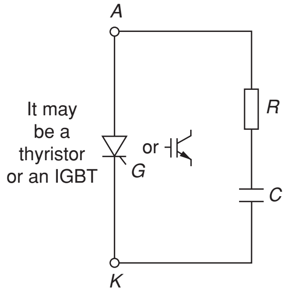



**Figure 6.37** Use of a snubber circuit across a power switching device

the circuit will discharge into this capacitor and charge the same to its optimum level (charging time constant $\tau = RC$) and slow down the rate of rise of TRV (r.r.r.v.), i.e. d$v$/d$t$ across the static circuit and limit the voltage spikes, similar to motor surge protection discussed in Section 17.10.1. The higher the value of $C$, the lower will be the voltage (commutation) overshoots. During a switch ON the capacitor discharges its total energy into the $R$ and prepares for the next switching operation. The power dissipation into $R$ is proportional to the switching frequency. $R$ also limits the peak value of the discharge current through the static device and damps the oscillations. Here the use of $C$ is to hold the charge and then release the same into $R$ and not to smooth the ripples.

- ***Current transients*** A similar situation will arise when a switching ON operation of the rectifier unit occurs when it is a thyristor rectifier. Under load conditions, the stored magnetic energy in the incoming supply system, which can be the feeding transformer and the line reactances similar to a fault condition discussed earlier, may cause a current transient which can be expressed by

$$V = -L \cdot \frac{di}{dt}$$

where
$V$ = applied voltage
$L$ = inductance of the total circuit up to the d.c. link and
$di/dt$ = rate of change of current, as the switching ON is a transient condition and causes over-load and short-circuits. This is maximum at the commencement of switching ON and becomes zero on its completion. It is analogous to contact making in an interrupter (Section 19.1.1). The same situation will arise even during a fault condition. Excessive rate of change of current may cause an over-load and even a short-circuit.

The rate of current change must therefore be controlled to a safe limit by providing a damping circuit on the supply side. This can be a series inductance as shown in Figures 6.26(a) and 6.34. This inductance may not be necessary when the unit is being supplied through a dedicated transformer. The inductor will absorb the magnetic charge and damp the rate of rise of current.

Now $V = L_i \dfrac{di}{dt}$

$L_i$ is the additional series reactance. The higher the value of $L_i$, the lower will be the rate of rise of current.

The inductor on the input side also suppresses the harmonics in the incoming supply, as high $L_i$ will provide a high impedance path to higher harmonics. For suppression of harmonics, where the supply system is already substantially distorted, additional $L$–$C$ filter circuits may be provided on the incoming side for more prominent harmonics. (For details of filter circuits see Section 23.9.) The main purpose of inductance here is protection, rather than suppression of harmonics.

Due to a high time constant of the damping circuit $\tau = L_i/R$ ($R$ being the resistance of the circuit) it will also delay occurrence of the fault by which time the circuit's protective scheme may initiate operation. It also adds to the line impedance to contain the severity of the fault conditions.

From the above we notice that the current surges can be caused either by the tripping of a current limiting device, when the distribution is through a large transformer on which is connected the static circuits, or by switching of the SCRs within the converter circuit itself. The protective scheme for both remains the same and is located at the incoming of the semiconductor circuits. There can be two situations. When the static circuits are being fed through a dedicated transformer in all probability no additional inductor will be necessary. Not even when there is a large transformer feeding a large distribution network on which is connected the semiconductor circuits. It is, however, advisable to carry out the trapped energy calculations to ensure that there is adequate inductance already available in the switching circuits of the semiconductor devices.

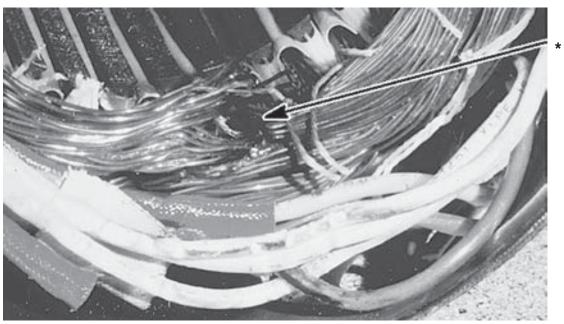
When there is no dedicated transformer and these circuits are connected on the system bus directly a large inductor will be essential at the incoming of the static circuits, sufficient to absorb the trapped charge within the transformer and the inter-connecting cables up to the converter unit. The size of the inductor can be calculated depending on the size (kVA) of the distribution transformer, its fault level and the characteristics of its current limiting protective device. An inductor sufficient to absorb $i_{st}^2 \cdot L$ of the transformer and the cables may be provided at the incoming of the static circuits.

### *Voltage surges in the inverter circuit*

Generally, voltage surges on an *LV* system are of little relevance as analysed in Section 17.7.6. Instances can, however, be cited of motor insulation failures, even on an *LV* system, when the machine was being controlled through a static drive, which may be an IGBT switched or a thyristor (GTO) switched inverter. The reason being a steep rising switching wave generated through the inverter circuit. The output of the inverter unit being in the shape of a non-sinusoidal voltage waveform also adds to the switching transients. To visualize the effects of fast switching in a static circuit, it is relevant to corroborate these with the switching of a conventional HV interrupting device, discussed in Section 17.7. The static devices also cause switching surges and their severity is also defined by their amplitude and the rise time (Figure 17.2). These devices are seen to produce voltage surges with an amplitude up to two to three times the voltage of the d.c. link and a rise time as low as 0.05–0.4 $\mu$s (typically) in IGBTs and 2 to 4 $\mu$s (typically) in GTOs (see Lawrence et al. (1996) and the Further Reading at the end of the chapter).

Drive manufacturers usually prefer using high switching frequencies as a measure to reduce harmonics and motor noise level but it gives rise to steep fronted waves. Higher the switching frequency shorter becomes the rise time ($t_1$) and steeper becomes the travelling wave (r.r.v = $V_t/t_1$ becomes very high) (see Figure 17.3). For significance of rise time see Section 17.6.2. Their severity increases when they cause a reflection (doubling phenomenon) at the motor terminals. The amplitude of the reflected wave will depend upon the length of the inter-connecting cables, between the inverter and the machine besides the switching frequency of the inverter unit. The amplitude of the wave after reflection may exceed the BIL (basic insulation level, Table 11.4 for LV and Table 11.6 for MV motors) of the machine, particularly when the machine is old and its insulation has deteriorated (Section 9.2). One such failure as a consequence of fast rising wave is detected in Figure 6.37(a). It is observed that this voltage may rise up to twice the d.c. link voltage or 3.1 Vr with a rise time as short as $t_1 \geq 0.1\mu$s. In IGBTs particularly, the rise time is too short and the surges behave like an FOW (front of wave, Section 17.3.3), which are all the more dangerous for the end turns of the connected machine. The length of cable from the inverter to the machine plays a vital role. The longer the cable, the higher will be the amplitude of the voltage surge at the motor terminals. This aspect is discussed in greater detail in Section 18.6.2. For example, for a switching surge with a wave front of 0.1 μs and prospective amplitude as 3.1 Vr, propagating at about 60 m/μs in the cable connecting the motor (Section 17.6.6), the safe cable length to arrest the wave reaching its first peak (3.1Vr) by the time it reaches the motor terminals should be less than $0.1 \times 60$ or $< 6$ m.

Too long a cable (multiple of safe length) will not be




*Failure of insulation usually in the first turn near the lamination stack.

**Figure 6.37(a)** Typical inverter spike damage to motor windings

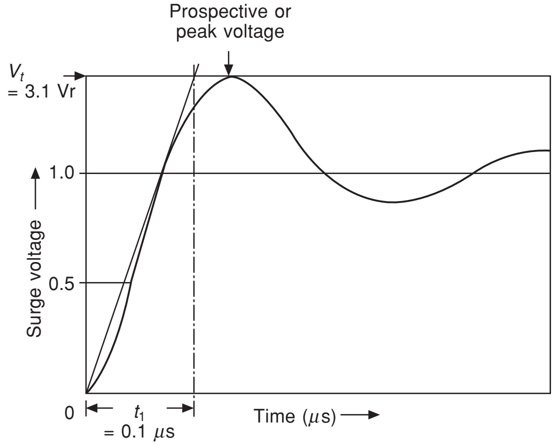



**Figure 6.37(b)** Severity of a surge damps as it propagates

safe to use in this case as the subsequent transient peaks can be more severe because of continuous switchings in the PWM inverter circuit. The surge severity surely damps as it propagates (Figure 6.37(b)) because of cable surge impedance but now the situation is more severe because of continuous switchings than discussed in Section 17.6.2.

The leading manufacturers of static drives specify, as a standard practice, the amplitude and the rise time of the switching surges of their devices at a particular voltage and switching frequency of their inverter unit (normally 2–16 kHz depending upon the rating of the machine as noted in Section 6.9.3). Also the maximum safe cable lengths (similar to HV interrupters (Section 17.8.1)). While most installations may not need separate surge protection, it is advisable to take precautions against any contingency during actual operation to protect the machine against these surges under the most onerous operating conditions.

## Derating of motors using static drives

– Insulation: As per IEC 60034-17 (also MG-I part 30) motors rated up to 500 V are usually capable to withstand such contingencies but motors above 500 V and up to 690 V may call for enhanced insulation system (usually termed as inverter duty insulation[a]) if adequate preventive measures as noted below are not taken. Nevertheless it is advisable to choose motors with enhanced insulation system only at least for new installations, to take care of any eventuality and ageing.

[a] Inverter duty insulation: It would mean better slot insulation, phase separators (Figure 6.37(c)) and reinforcement and bracing of overhangs etc. Besides better impregnation with additional immersions of the whole stator in the insulating varnish similar to power house treatment (Section 9.3.1). It is noticed that sometimes in absence of an inverter duty motor being readily available the user may instead choose for a higher class of insulation motor such as class F or H. This measure too can bear the severity of surges and harmonics to some extent and also enhance the rating of the machine. It is, however, advisable to use an inverter compatible motor as far as possible and check the safe cable length with the drive supplier and take extra precautions if the actual length exceeds the safe length.

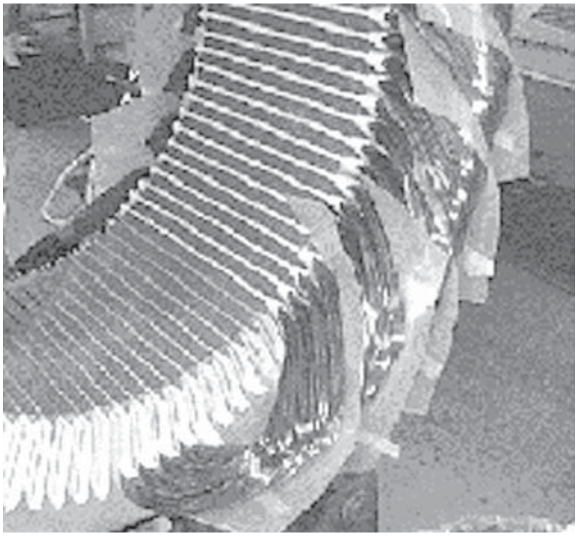


**Figure 6.37(c)** Insulation separation of end coils

For existing installations undergoing retrofitting with inverter drives (to have smoother control of motors and to also conserve on energy such as in case of pumps, compressors and fans) it is advisable to retrofit the old motors with EE motors suitable for inverter duty (enhanced insulation system). And where it may be onerous to change the motor, measures may be taken through $dv/dt$ and $di/dt$ controls to limit the severity of switching surges on the load side as noted next.

– Rating: It may be checked on case to case basis. Where the variation in speed is by virtue of load demand itself such as in pumps, fans and compressors the normal inverter duty motor would suffice, as the drive is primarily for energy conservation. But where the motor is required to perform specific duty at constant torque at lower speeds (Figure 6.7(b)), the derating of the motor must be considered for poor cooling, lower efficiency and extra heating due to harmonics[b], besides the cable length factor. Some consultants consider this derating as 10%. It will however, be more prudent to do an actual analysis to arrive at the optimum size of motor to save on unwanted extra rating and consequent higher cost of drive. One must however choose for EE and inverter duty motors as far as possible.

## Remedies from harmonics, over-voltages and voltage surges

The remedies are the same as those discussed in Section 17.10 on the protection of electric motors. For example, surge capacitors for most of the motors will prove sufficient and economical to protect the machine by taming the steepness of fast rising surges at the motor terminals. (See curves 1 and 2 of Figure 17.21.) The surge capacitor is provided with a discharge resistance as standard practice, as also noted in the snubber circuits. The resistance would help the capacitor discharge quickly and prepare for the next operation. The load side steep rising wave can also be tamed by

i. **Load side reactor** (to tame $di/dt$) (Figure 6.37(d)) These are specially designed reactors to reduce $di/dt$, i.e. rate of rise of current surge and in turn $dv/dt$. The value of $L_1$ is decided as per cable length such that by the time the wave reaches the motor terminals the transient voltage rise is within safe limits. As a standard practice drive manufacturers furnish different cable lengths for their drives and one can choose a drive suiting his cable length.
ii. **Load side filters** (to tame $dv/dt$) (Figure 6.37(d)) Now *L-C* filter circuits are used to damp $dv/dt$ in both magnitude and steepness typically to < 500 V/µs.
iii. **Sinusoidal filters** They are also filter circuits like the above but more closely tuned to suppress voltage over-shoots and provide a near sinusoidal waveform at the motor terminals. It is a costly arrangement.

*Note*

A manufacturer of static drives would usually give an option to the user to operate their inverter circuit (IGBTs or IGCTs normally) at high PWM carrier frequencies (typically 2–16 kHz) to smooth the output (load side) voltage. But at high frequencies, the propagation of surges becomes faster and may cause quick reflections, which would require either a shorter cable or the use of a surge suppressor. High-frequency operations also raise the noise level in the ground

[b] Harmonics cause extra $I^2R$ stator and rotor copper and iron losses.

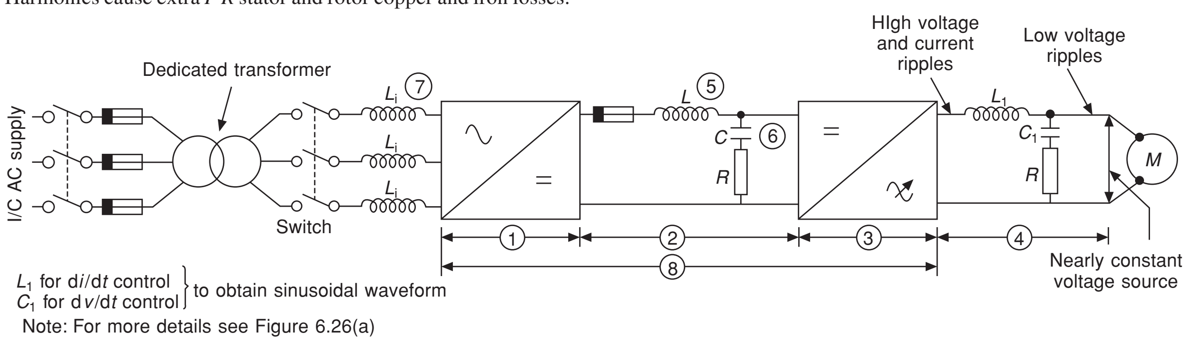



**Figure 6.37(d)** Taming the PWM inverter output waveforms

path and can cause sensitive devices like PLCs, sensors and analogue circuits to behave erratically, as they are all connected through the ground circuit. It is therefore desirable to operate the IGBTs/IGCTs or SCRs/GTOs at lower frequencies, preferably 2–6 kHz, as this will cause low ripples as well as a low noise level. A moderate carrier frequency will also help in taming the arriving surges at motor terminals with only moderate steepness.

### 6.13.3 Grounding practices and shielding of signals in electronic circuits [For EMC/EMI requirements see Section 23.18]

An electronic circuit may be associated with many in-house services such as management information services (MIS), data transfer for a process, system automation as for a SCADA system (Section 24.11) and required to perform the following,

– Sensing of operating conditions
– Voice mail communication and serial data transfer to peripherals.
– Services within cabinets and
– Services between cabinets etc

These services, usually part of a power system, are subject to strong EM disturbances. To make the power information and data transfer system free from EM effects it would call for meticulous

(i) Grounding System – that should be an isolated or a clean grounding system, and
(ii) Shielding of data and signals in cables before grounding them.

#### (i) Grounding System

Purpose
– Not only safety to personnel by limiting the touch voltage to 50 V but also to possess reliability, integrity and noise control (EMI effects) for sensitive electronic equipment and devices connected to it, to discharge the above services reliably. The maximum EMI problems in the form of noise 'humming' originate from ground loop and that is due to potential ingredients between two safety grounds separated by a large distance.
– Reduce common mode voltage between neutral and ground. It should be less than 0.5 V. Higher voltage would mean system lock-up, communication distortions, corrupted or unreliable data even danger to equipment.
– Handle high harmonic voltages and currents.
– Provide safe path for the discharge of ground faults and lightning strokes.
– Provide safe path through the cable shields to electrostatic discharges in case of high harmonic electronic circuits, particularly PWM inverter fed loads. Like providing path to bearing currents (Sections 6.14 and 10.4.5).
– It is of utmost importance to save the signals from drive to equipment, shaft currents and the communication network.
– Provide safe path to EMI and RFI waves and comply with EMI/EMC requirements.
– Isolate sensitive equipment from inductive loads (inductive loads generate high EM disturbances).
– Eliminate shared circuit equipment problems (unbalanced loads with shared neutrals).
– Eliminate noise by separating case (electronic circuits) and equipment grounds. Also providing an effective low resistance safe path to electronic data and signal shielding. Ground to eliminate noise means an equi-potential ground system as noted above.

Isolated star ground system is the most preferred system for electronic circuits to eliminate neutral circuit and hence, the voltage ingredient between neutral and ground. While it may not be the best grounding procedure as noted in Section 20.3, it provides the least compromise in case of electronic circuit grounding. This is now the latest approach to controlling noise in audio systems to comply with EMI/EMC norms as per Audio Engineering Society (AES), USA.

***Isolated ground (IG) or Clean ground***
(see also IEC 61000-5-2 and IEEE 142)

Isolated ground (IG) or clean ground is a technique for sensitive electronic circuits (ECs) to reduce electromagnetic (EM) and radio frequency (RF) interferences and provide a noise-free system besides preventing personnel from electric shocks. IG insulates the ground of sensitive electronic circuits from the electrical equipment ground system. Thus, the ground potential shift due to stray ground currents flowing in the electronic grounding circuit is eliminated and provides the required EMI/RFI shielding. It may clearly be noted that electronic equipment grounding circuit is not a separate ground as it also ultimately gets connected to the same main ground. In fact there cannot be more than one grounding system for one installation. We can instead say that there are two isolated ground paths terminating at the same (main) ground. A separate ground system isolated from power ground is not only ineffective and unsafe it can even cause more noise, because of potential ingredients between the two grounds.

*Note*
– In a separate grounding system potential difference between different ground locations such as between electronic grounding loop and building main ground may cause circulating currents through data cables and computer printer cables and become a source of noise and EMI disturbances to all electronic devices and gadgets connected to it.
– Other than line disturbance or fault condition difference in potential can also be a result of bad joints or bonding.

For easy identification this kind of IG system is provided with a colour code, usually.
• Ground circuit for clean ground – Green (according to IEEE 142 green with yellow strip)
(IEC 60439-1 specifies twin coloured, green and yellow)
[To identify clean ground circuit the IG receptacles are provided with an orange triangle on their faces as a caution not to bond them with the main ground circuit.]
• Main ground wire – Bare copper (green*)
• Case/body neutral – White or yellow (white*)

- Phase wires – red, yellow, blue (black*)
  *According to IEEE 142

### (ii) Shielding of signals

***Routing and segregation of cables***

Routing, segregation, separation distances and directions of cables are important parameters to provide a desirable degree of shielding, e.g.

- Wiring between the main input and an RFI filter must be separated by a minimum of 300 mm. So also it must be separated from all other drives and drive output components (enclosures, modules, inductors and motor cables). Electrical separation is a means to minimize proximity effect. For details on proximity effect see Section 28.8.4.
- Control and power wiring must cross at 90° and be also separated by a minimum of 300 mm.
- If an EMC filter is fitted, cables on the input side of the filter must be separated as above from cables and components on the drive side of the filter.
- Ribbon cables should be run along the grounded metal and not through the air. If ribbons are too long they may be folded rather than rolled to nullify the EM distortions.
- The screen of screened cables must be bonded to ground at both ends at the main ground point.
- Where the cable of electronic circuits enters or leaves an enclosure it should be bonded with its own isolated ground.

An electronic equipment manufacturer would know the potential locations that may give rise to ground voltages and electrostatic leakage currents that cause noise. For instance high frequency capacitors usually leakage capacitances that discharge at high frequencies, or high value (kVAr) electrostatic capacitors in electric power circuits cause electrostatic leakage currents. Such as on the load side of a PWM inverter there are electrostatic discharges in a motor (Sections 6.14 and 10.4.5). Even d.c. link during switching operations causes heavy leakage currents. All these electrostatic discharges are potential sources of ground voltages and responsible for noise.

Manufacturers take preventive measures to eliminate the same at source by proper shielding and grounding of cables. Brief philosophy of shielding of signals is presented in Section 13.3.6 and Figure 13.13. Below we provide some vital information about shielding of various kinds of electronic signals needing integrity and reliability.

***EM compatible control cables***

It is important to use correct types of control cables for electronic circuits to comply with EMC/EMI requirements

*Analogue Signals* – Each analogue signal must use a separate double screened twisted pair of cables. A common return path will corrupt the signals.

*Digital Signals* (110/230 V a.c.) – Cables as for analogue signals, but now an unscreened multi-core cable can also be used. But a.c. and d.c. control voltages should not be mixed. There must be separate cables for a.c. and d.c. voltages.

*Digital signals (24V d.c.)* – Cables as for analogue signals but now a single screen twisted and multi-core cable is permissible.

*TG analogue or pulse encoder cables* – They should be twin twisted pair with an overall screen. The screen should remain continuous up to the encoder. If it needs to be broken such as when entering or leaving a junction box or metal box (usually Faraday cage), the integrity of the screen must be maintained by enclosing the terminations inside the box in such a fashion that it makes close bonding as if within a Faraday cage. The screen of the encoder cable also shall terminate to the main ground terminals at both ends. The length of the cable should be restricted (say within 15 m) to limit the voltage drop to maintain the necessary encoder voltage at the encoder terminals.

*Communication cables* – The cables recommended for a particular communication standard should only be used and installed as recommended by the appropriate 'body'. Such as in case of DeviceNet, a deviceNet recommended cable such as Belden 3084 A must be used and installed as recommended by Open DeviceNet Vendors Association (ODVA)*.

## 6.14 PWM inverter drives causing shaft currents

When a motor is driven through a PWM inverter unit, a peculiar phenomenon of bearing currents and their failure as a consequence of it is noticed in certain range of motors. Typically for frame sizes ≥315 and inverter switching frequency $fs$ >10 kHz. The situation becoming worse at higher frequencies (IEC 60034-17 and ref. 11 under Further Reading).

The windings leakage capacitances (Figure 10.6(a)) at high pulses of the inverter output voltage offer a very low reactance ($X_c \propto 1/fs$) and may discharge even at low stator voltages. The effect of high frequency now is similar to, rather worse than that of high voltages in MV motors noted in Section 10.4.5 and induces shaft voltages as a result of capacitive coupling. In fact the very high frequencies cause much higher shaft voltages and consequent much larger leakage currents than in MV motors. These shaft voltages are noticed up to 10–30 V (for frame sizes above 315). At high frequencies (>10 kHz) the grease dielectric may break down at about 10–15 V and allow the leakage current flow through the bearings. As a result bearings emanate sort of a scratching noise. These currents if not prevented are dangerous and may cause bearings' failure. Prevention from shaft currents

* ODVA is an international organization that supports network technologies built on the Common Industrial Protocols (CIP), such as DeviceNet, EtherNet/P, CIP syn and CIP safety. CIP allows companies to integrate I/O control, device configuration and data collection across multiple networks.

in PWM inverter drives is similar to as discussed in Section 10.4.5.

## 6.15 Energy conservation using static drives

### *While the motor is operating under-loaded*

In the various types of static drives discussed so far, the supply voltage would adjust automatically at a level just sufficient to drive the motor to meet its load requirements. Hence it is not necessary for the motor to be applied with full voltage at all times. The voltage adjusts with the load. This is an in-built ability of a static drive that would save energy and losses. One would appreciate that most of the industrial applications consider a number of deratings and safety margins while selecting the size of the motor to cope with a number of unforeseen unfavourable operating conditions occurring at the same time. This is discussed in Chapter 7 (see also Example 7.1). The size of the motor is therefore chosen a little larger than actually required. As a consequence even when the motor may be performing its optimum duty, it may hardly be loaded by 60–80% of its actual rating causing energy waste by extra iron and copper losses and operating at a reduced p.f. Static drives are therefore tangible means to conserve on such an energy waste.

### *While performing a speed control*

A very important feature of solid-state technology is energy conservation in the process of speed control. The slip losses that appear in the rotor circuit are now totally eliminated. With the application of this technology, we can change the characteristics of the motor so that the voltage and frequency are set at values just sufficient to meet the speed and power requirements of the load. The power drawn from the mains is completely utilized in doing useful work rather than appearing as stator losses, rotor slip losses or external resistance losses of the rotor circuit. Many industries use fixed speed motors for variable speed requirements through gears or belt-drives. These devices add to losses and the motor may also be running under-loaded on many occasions. This is a colossal drain of useful energy and must be saved at all costs (Section 1.19). With the thrust on energy conservation one must use energy efficient (EE) motors and static drives and adapt all such practices that minimize losses and conserve energy as far as possible. More so when the payback period is so short.

### 6.15.1 Illustration of energy conservation

In an industry there may be many drives that may not be required to operate at their optimum capacity at all times. The process requirement may require a varying utilization of the capacity of the drive at different times. In an induction motor, which is a constant speed prime-mover, such a variation is conventionally achieved by throttling the flow valves or by employing dampers.

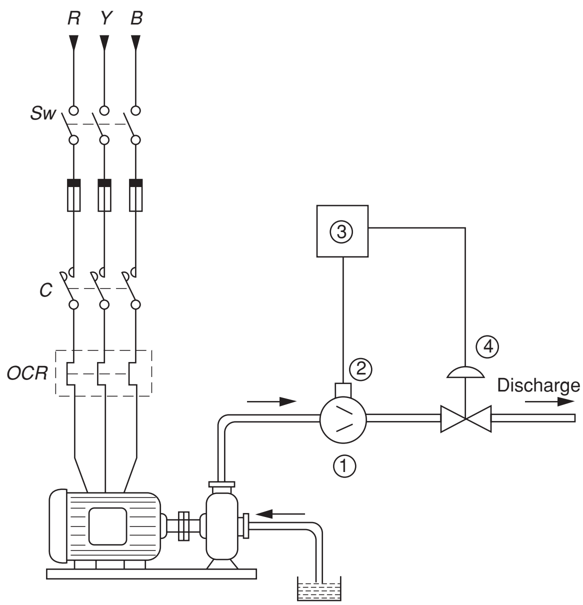
There may be many types of drives in an industry, particularly when it is a process industry. The most common drives are fans, pumps, and compressors etc., employed for the various utilities, storage and process activities of the plant. The plant may be chemical or a petrochemical, water treatment or sewage disposal, paper and pulp unit or even a crane or a hoist application.

The method of speed or flow control by throttling, damping (vane control) or braking, indirectly reduces the capacity of the motor at the cost of high power loss in the stator and slip loss in the rotor circuit, as discussed above. These losses can now be eliminated with the effective use of static control variable-speed drives or fluid couplings. We will show, through the following illustrations, the energy saving by using such controls.

### *Throttle, damping or vane control*

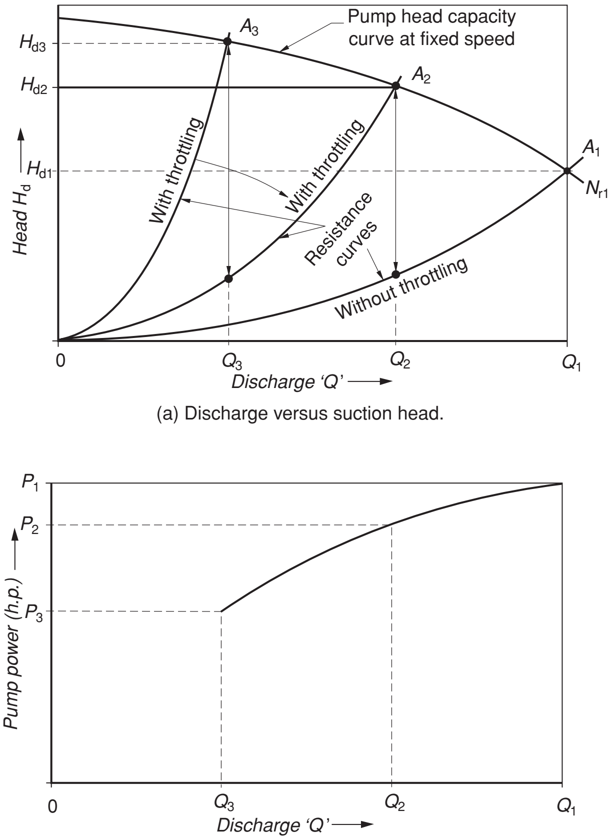
For ease of illustration we will consider the characteristics and behaviour of a centrifugal pump which is similar in behaviour to radial/axial flow fans and centrifugal/screw compressors. Figure 6.38 shows the mechanical connection of a flow valve to control the output of the pump or the discharge of the fluid through the throttle of the valve. Figure 6.39 illustrates the characteristics of the pump:

- Discharge versus suction head, i.e. $Q$ versus $H_d$ and
- Discharge versus pump power requirement, i.e. $Q$ versus h.p.




① Venturimeter – to measure the velocity of fluid
② Probe to sense the velocity of fluid
③ Flow meter or sensor – To convert the velocity of fluid to the rate of flow
④ Motorized sluice valve – To throttle the flow of fluid

**Figure 6.38** Conventional throttle control




(a) Discharge versus suction head.

(b) Discharge versus pump power.

**Figure 6.39** Power requirement and rate of discharge on a throttle control

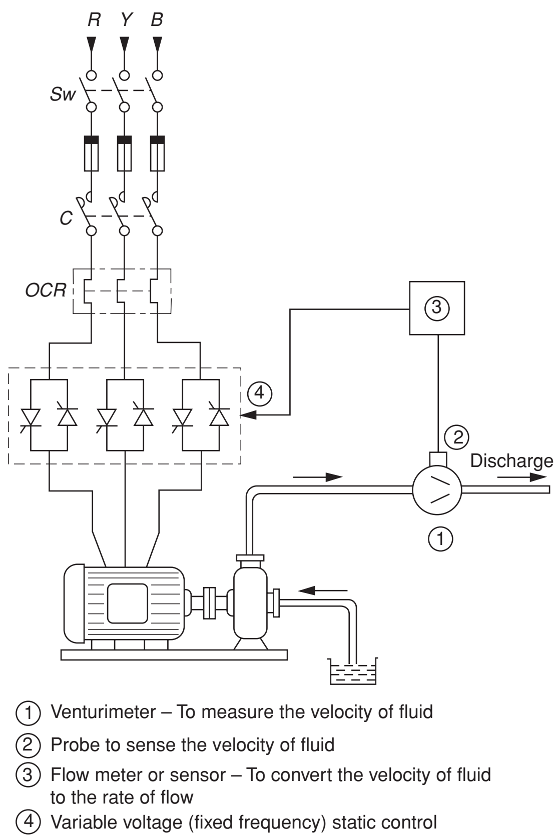



**Figure 6.40** Energy conservation through static control

The rated discharge is $Q_1$ at a static head of $H_{d1}$ and a motor h.p. $P_1$. In the process of controlling the discharge from $Q_1$ to $Q_2$ and $Q_3$, the valve is throttled, which increases the head loss of the system (or system resistance) from $H_{d1}$ to $H_{d2}$ and $H_{d3}$ respectively. The operating point on the $Q$–$H_d$ curve now shifts from point $A_1$ to $A_2$ and $A_3$ as a result of back pressure. The pump power requirement now changes from $P_1$ to $P_2$ and $P_3$ on the $Q$–h.p. curve. We can see that, due to added resistance in the system, while the discharge reduces, the corresponding power requirement does not reduce in the same proportion.

### *Flow control through static control*

The same operating control, when achieved through the use of a solid-state control system will change the mechanical system to that of Figure 6.40. We have used a simple, full-wave, phase-controlled, variable-voltage solid-state device, employing a triac (two SCRs in anti-parallel). The voltage to the motor is monitored through a flow sensor, which converts the flow of discharge through a venturimeter to electrical signals. These signals control the voltage and adjust the speed of the motor to maintain a predefined discharge flow. The use of a throttle valve is eliminated, which in turn eliminates the extra head loss or system resistance. Figure 6.41 illustrates the corresponding characteristics of the pump with this type of flow/speed control.

To reduce the discharge from $Q_1$ to $Q_2$ and $Q_3$ in this case, the speed of the pump, and so also of the motor, reduces from $N_{r1}$ to $N_{r2}$ and $N_{r3}$. The $Q$–$H_d$ characteristics change according to curves $N_{r2}$ and $N_{r3}$, at a corresponding pump power requirement of $P_2'$ and $P_3'$ respectively, according to power curve $P'$. These power requirements are significantly below the values of $P_2$ and $P_3$ of Figure 6.39 when discharge control was achieved by the throttle. The system resistance curve remains unaffected, whereas the pump power demand curve traverses a low profile as in curve $P'$ due to lower speeds $N_{r2}$ and $N_{r3}$. The power requirement diminishes directly with speed in such pumps.

The energy saving with this method is considerable, compared to use of the throttle, which is also evident from curve $P'$ of Figure 6.41.

## 6.15.2 Computation of energy saving

Consider Figure 6.42 with typical $Q$–$H_d$ curves at different speeds and different system resistances, introduced by

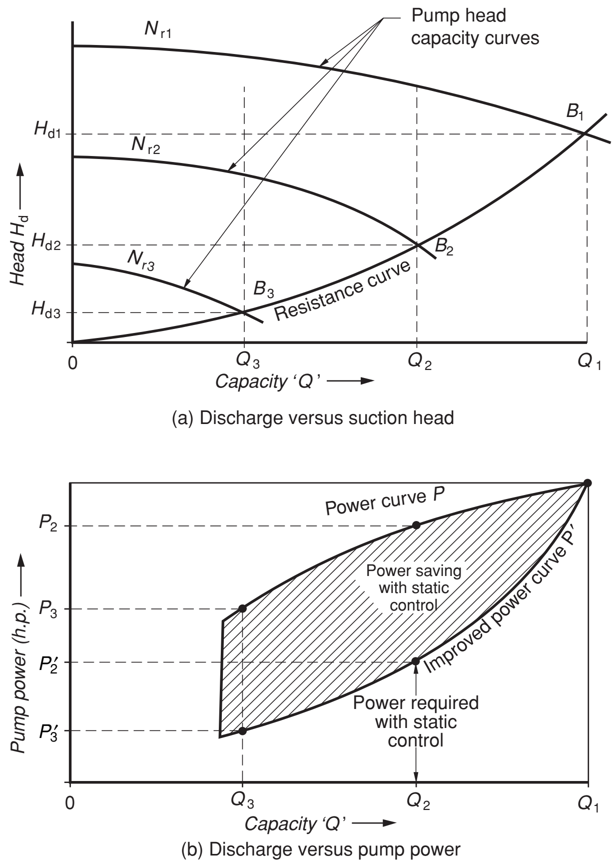



**Figure 6.41** Power saving with the use of static control compared to throttle control

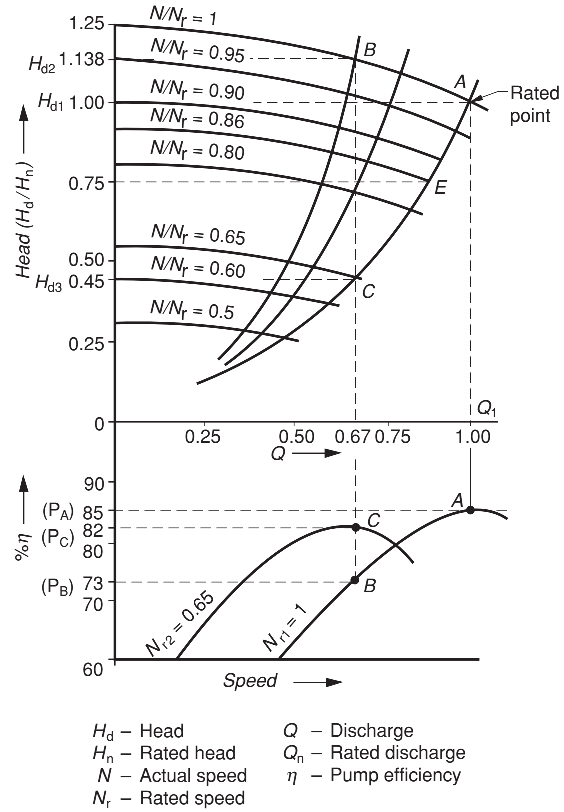



**Figure 6.42** Discharge versus head curves of a pump at different speeds and resistances introduced by throttle

the throttle. Point $A$ refers to the rated discharge $Q_1$ at rated speed $N_{r1}$ and head $H_{d1}$ when the throttle valve is fully open. Lets us consider the condition when the discharge is to be reduced to say, $0.67Q_1$.

### *Throttle control*

The system resistance increases and discharge reduces at the same rated speed $N_{r1}$. This condition refers to point $B$, to which the earlier point $A$, has now shifted. The system now operates at a higher head $H_{d2}$, whereas the actual head has not increased. This condition has occurred due to higher system resistance offered by the throttle. The pump and the prime-mover efficiency will now reduce to 73% from its original 85%.

### *Static control*

The same discharge of 0.67 $Q_1$ is now obtained without throttling the valve, but by controlling the speed of the motor to 0.65 $N_{r1}$. Point $A$ now shifts to point $C$ on the $Q$–$H_d$ curve for 0.65 $N_{r1}$. For this reduced discharge of 0.67 $Q_1$ the suction head $H_d$ also reduces to 0.45 $H_{d1}$ and so also does the system resistance. The efficiency of the system is now set at 82%.

There is thus an obvious energy saving by employing static control over the conventional and more energy consuming throttle control.

### *Extent of energy saving*

**Example 6.1**

This can be determined by

$$P = \frac{H_d \cdot Q \cdot d}{36 \cdot \eta} \qquad (6.7)$$

where

$P$ = shaft input in kW
$H_d$ = head in bar
$Q$ = discharge in m$^3$/hour
$d$ = specific gravity of the liquid in gm/cm$^3$
$\eta$ = efficiency of the pump

(i) Rated power required by the pump

$$P = \frac{H_d \cdot Q \cdot d}{36 \times 0.85}$$

(ii) Power required when discharge is controlled through the throttle,

head $H_{d_2} = 1.138\ H_{d_1}$
discharge = 0.67 $Q$
$\eta$ = 0.73

$$\therefore \quad \text{Power} = \frac{1.138\ H_{d_1} \times 0.67\ Q \cdot d}{36 \times 0.73}$$

$$\therefore \quad \frac{\text{Ratio of power through throttle}}{\text{Rated power required}}$$

$$= \frac{1.138 \times 0.67}{0.73} \times 0.85$$

$$\simeq 0.89$$

i.e. the power reduction with this method is around 11% of the rated power requirement.

(iii) Power required when discharge is controlled through static control, i.e. by speed variation;

Head $H_{d_3} = 0.45\ H_{d_1}$

discharge = 0.67 $Q$

$\eta = 0.82$

$$\therefore \quad \text{Power } \frac{0.45\ H_{d_1} \times 0.67\ Q \cdot d}{36 \times 0.82}$$

$$\therefore \quad \frac{\text{Ratio of power through speed variation}}{\text{Rated power required}}$$

$$= \frac{0.45 \times 0.67}{0.82} \times 0.85$$

$$\simeq 0.31$$

i.e. the power reduction through speed control is 69% that of the rated power requirement.

Thus the energy saving in this particular case by employing the method of speed control over that of throttle control will be

= 69% – 11% = 58%

**Example 6.2**
To determine saving of costs through speed control, consider the following parameters, for the above case:

$P$ = 100 kW

Discharge = 67% of the rated flow
Duration of operation at reduced capacity = say, for 25% of total working hours.

Energy tariff = say, Rs.1.2 per kWh

∴ Total energy consumed per year, considering 300 operating days/year and 24 hrs/day

= 100 × 24 × 300

= 720 000 kWh

Energy consumed while operating at the reduced capacity for 25% duration = 0.25 × 720 000

= 180 000 kWh

∴ Energy saving = 0.58 × 180 000

= 104 400 kWh

And total saving in terms of cost = 1.2 × 104 400

= Rs.125 280 per year

**Example 6.3**
**Energy saving in a slip recovery system**
Consider an I.D. fan of 750 kW, running at 75% of $N_r$ for at least 20% of the day. Considering the load characteristic in cubic ratio of speed,

∴ $P$ at 75% speed

= $(0.75)^3 \times 750$ kW

≃ 316 kW

and slip power,

$P_s$ = (1 – 0.75) × 316 kW

= 0.25 × 316

= 79 kW

Considering the efficiency of slip recovery system as 95% and 300 operating days in a year, the power feedback to the main supply source through the slip recovery system in 20% of the time,

= [0.95 × 79] · [0.2 × 300 × 24]

= 108 072 kWh

And in terms of costs at a tariff rate of Rs 1.2 per unit,

= Rs 1.2 × 108 072

= Rs 129 686.4 per year

## 6.16 Application of static controllers

### 6.16.1 Soft starting (reduced voltage starting)

This facilitates a stepless reduced voltage start and smooth acceleration through control of the stator voltage. The smooth voltage control limits the starting torque ($T_{st}$) and also the starting inrush current ($I_{st}$) as illustrated in Figure 6.43. This is achieved by arranging a full-wave phase-controlled circuit by connecting two SCRs in anti-parallel in each phase as illustrated in Figure 6.44. Varying the firing angle and the applied voltage will also affect the dynamic phase balancing in each phase, and control the $I_{st}$ and $T_{st}$ as programmed. Increasing the firing angle decreases the angle of conduction and causes the voltage to decrease, as seen in Figure 6.23. The starting voltage, i.e. the angle of firing, is pre-set according to the minimum voltage required, to ensure a desired minimum $T_{st}$ and permissible maximum $I_{st}$ which can be predetermined with the help of the motor characteristics and the load requirements. The voltage is raised to full, gradually and smoothly, but within a preset time, as determined by the motor and the load characteristics. The size of the static starter will depend upon the starting current chosen and the corresponding starting time. Generally, the normal practice of the various manufacturers, as noted earlier, is to define the size of their soft starters, based on a starting current of 150% of $I_r$ and a starting time of up to one minute. A higher starting current or a higher duration of start may call for a higher deration or larger starter. The starter therefore provides no control over the starting current, which is a function of the applied voltage.

It is also possible to perform a cyclic duty having some no-load or a light load and some fully loaded periods, as discussed in Section 3.3. The firing angles of the SCRs can also be programmed accordingly to reduce the applied voltage to the motor to a minimum possible level, during no-load or light-load periods, and hence conserving an otherwise wasted energy by saving on the no-load losses. Different mathematical algorithms are used to achieve the desired periodic $T$–$N$ characteristics of the machine.

During start-up, the firing angle is kept high to keep the $V$ and $I_{st}$ low. It is then reduced gradually to raise the

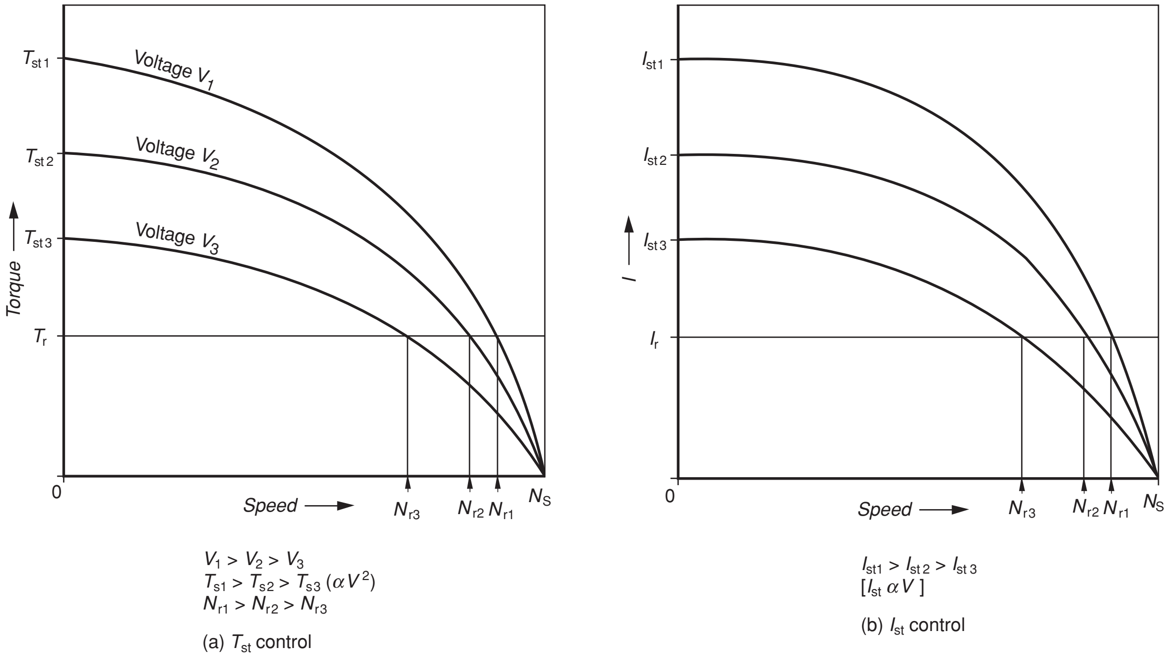



**Figure 6.43** Speed control by varying the applied voltage (use of higher-slip motors)

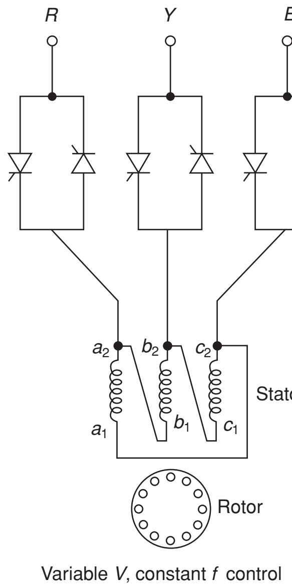



**Figure 6.44** Stator static voltage control (soft starting)

voltage to the required level. There is thus a controlled acceleration. The maximum $I_{st}$ can also be regulated to a required level in most cases, except where the $T_{st}$ requirement is high and needs a higher starting voltage, when it may not be possible to limit the $I_{st}$ to the desired level. The scheme is a simple reduced voltage start and does not facilitate any speed control. Since the starting current can now be regulated to a considerable extent, the scheme is termed a soft start. Now $T_{st} \propto V^2$ and $I_{st} \propto V$. This may be a good alternative to $Y/\Delta$ or autotransformer starting by eliminating current overshoots during the changeover from $Y$ to $\Delta$ in a $Y/\Delta$ or from one step to another in an autotransformer start. It also prevents an open transient condition (Section 4.2.2(a)). Also in a soft start the switching SCRs are bypassed through a contactor after performing a starting duty similar to a $Y/\Delta$, $A/T$ or a wound rotor start. In conventional startings also the star contactor, the autotransformer or the external resistances, respectively, are cut off from the circuit, after performing their starting duty. Such a provision protects the starter unit from avoidable voltage strains, saves heat losses and cools it to prepare it for the next switching operation. The starter is also free to perform switching duties on other machines of a similar size connected on the same line, if so desired, to save costs. The starter can be even used as a switching kit for a group of similar-sized motors.

The voltage can be varied from 0% to 100% and hence the $I_{st}$ can be limited to any required level. Such a linear voltage variation may be suitable in most cases, particularly when the motor has to pick-up a light load or is at no-load. Motors driving heavy inertia loads or loads requiring a high starting torque or both may not be able to pick-up smoothly with a linear voltage rise from 0 volts. This may cause a locked rotor or stalling condition until the voltage rises to a level sufficient to develop an accelerating torque, capable of picking up the load. It is also possible that it may need a prolonged starting time not commensurate with the thermal withstand time of the motor or a larger starter. In such cases, it is essential to have a minimum base or pedestal voltage as illustrated in Figure 6.45(a). The voltage is adjusted to the lowest possible level, so that the $I_{st}$ is kept as low as possible. In

Figure 6.45(b) we illustrate a motor with a normal starting current of 650% $I_r$ at the rated voltage. To limit this to a maximum of 350% of $I_r$, we have provided a base voltage of about 350/650 or 54% of the rated voltage. The minimum $T_{st}$ is, however, matched with the load requirement to attain the rated speed within its thermal withstand time. For more details see Section 3.5 and Example 7.1. The voltage is then raised so that during the pick-up period the $I_{st}$ is maintained constant at 350%, until the motor reaches its rated speed. As a usual practice these starters are provided with ramp-up and ramp-down facilities to meet such a situation.

Since it is not practical to custom-build each starter, the normal practice by manufacturers of soft starters is to provide a variety of motor parameters to which nearly all motors will fit and the user may select the motor that best suits the load requirements. Such starters from small to very large sizes (even up to 5 MW) are usually available off the shelf. And as noted before can be used as starting kits for many motors of similar ratings in the same plant.

The other advantages of a solid-state soft start are that an unbalanced power supply is transformed to a balanced source of supply automatically by suitably adjusting the firing angle of each SCR through their switching logistics. Also a low starting current can economize on the size of switchgears, cables and generator where a captive power is feeding the load. These starters are very cheap as there is no inverter or rectifier unit now.

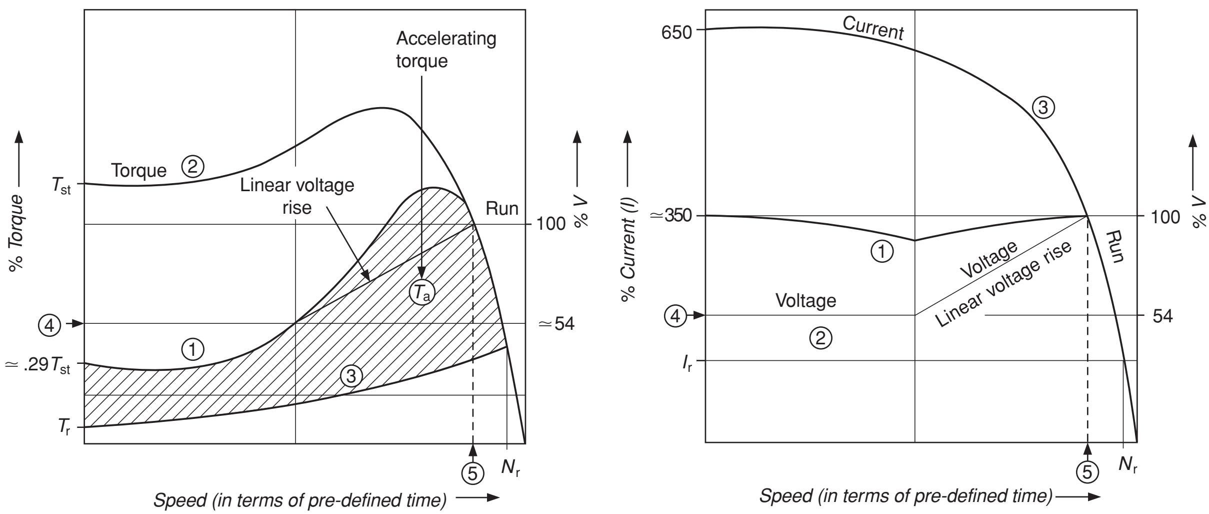
### 6.16.2 Soft stopping

The normal method to stop a motor is to do it instantly. This is accepted practice in most cases and the machine stops as shown in Equation (2.13). A higher moment of inertia (*MI*) of the driven masses will mean a reasonably longer duration to come to a standstill, while a low *MI* will mean a faster stop. But in loads requiring high braking torques, such as conveyor systems, escalators and hoists, the stopping time may be too short. In some cases, such as a pump duty, the stoppage may be near-abrupt. Such a situation is not desirable and may cause shocks to the motor. In a pump, an abrupt stoppage may cause severe shocks and hammering effects on pipelines, due to back-flow of the fluid. Shocks may even burst pipelines or reduce their life when such stoppages are frequent. They may also damage non-return valves and other components fitted on the lines and thus weaken the whole hydraulic system.

A gradually reducing voltage rather than an instant switch OFF is therefore desirable for all such applications. This would also gradually reduce the flow of the fluid, leading to a strain-free and smooth stopping of the machine. In such cases, a soft stop feature, similar to a soft start, can be introduced into the same starter, which would gradually reduce the stator voltage and facilitate a smooth and shock-free stop.

### 6.16.3 Slip recovery system (to control wound rotor motors)

As discussed earlier, the motor speed–torque characteristics depend largely upon the rotor current (Equation (1.1)) and rotor resistance (Equation (1.3)).




(1) Approximate torque curve during a soft start
(2) Normal torque curve
(3) Load torque
(4) Base or pedestal voltage
(5) Soft starter can be removed from the circuit and used to start other motors, if desired.

(a) Torque characteristics

(1) Approximate current curve during a soft start
(2) Voltage can be adjusted to maintain the starting current constant at 350% (or any desired value)
(3) Normal torque curve
(4) Base or pedestal voltage
(5) Soft starter can be removed from the circuit and used to start other motors, if desired.

(b) Current characteristics

**Figure 6.45** Current and corresponding torque characteristics of a motor during a soft start

The higher the rotor current or resistance, the higher will be the starting torque as illustrated in Figure 6.46, and the higher will be the slip and slip losses as well as reduced output. The maximum torque is obtained when $R_2$ and ${}_{ss}X_2$ are equal. Speed control can be achieved by varying the rotor resistance or by varying the rotor current $I_{rr}$. The slip recovery system provides an ideal control, employing a basic converter unit, supplemented by a vector or field oriented controlled inverter unit in the rotor circuit of the motor, as illustrated in Figure 6.47. The inverter unit controls the power flow from the rotor to the mains, thus acting as a variable resistance. The stator operates at a fixed frequency.

The inverter may be a current source inverter, rather than a voltage source inverter (Section 6.9.4) since it will be the rotor current $I_{rr}$ that is required to be varied (Equation (1.7)) to control the speed of a wound rotor motor, and this can be independently varied through the control of the rotor current. The speed and torque of the motor can be smoothly and steplessly controlled by this method, without any power loss. Figures 6.47 and 6.48

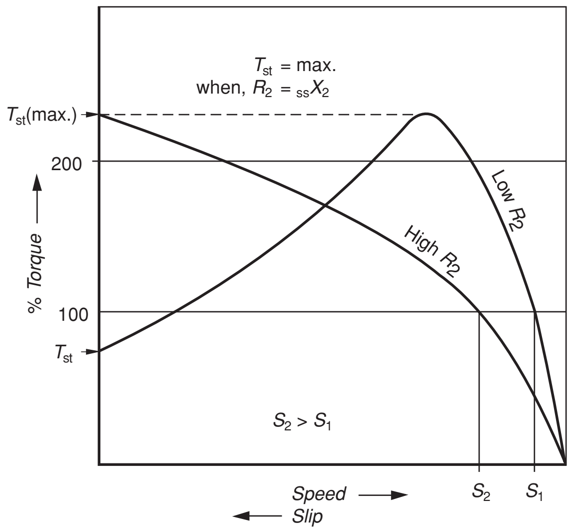



**Figure 6.46** Effect of rotor resistance on torque

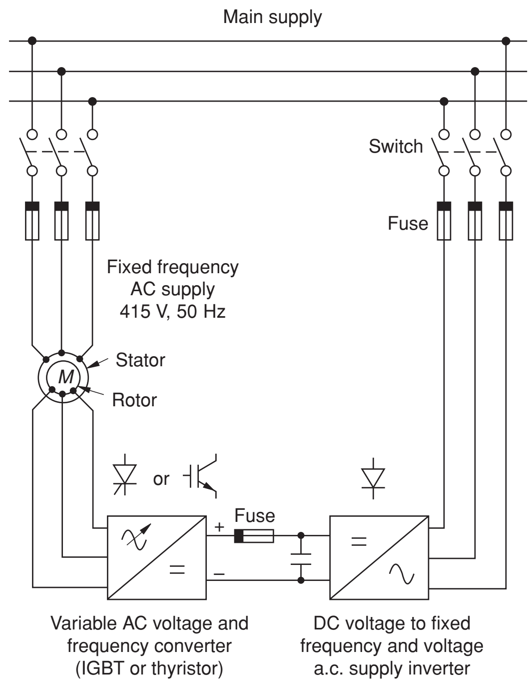



**Figure 6.47** Slip-ring motor control showing slip recovery system

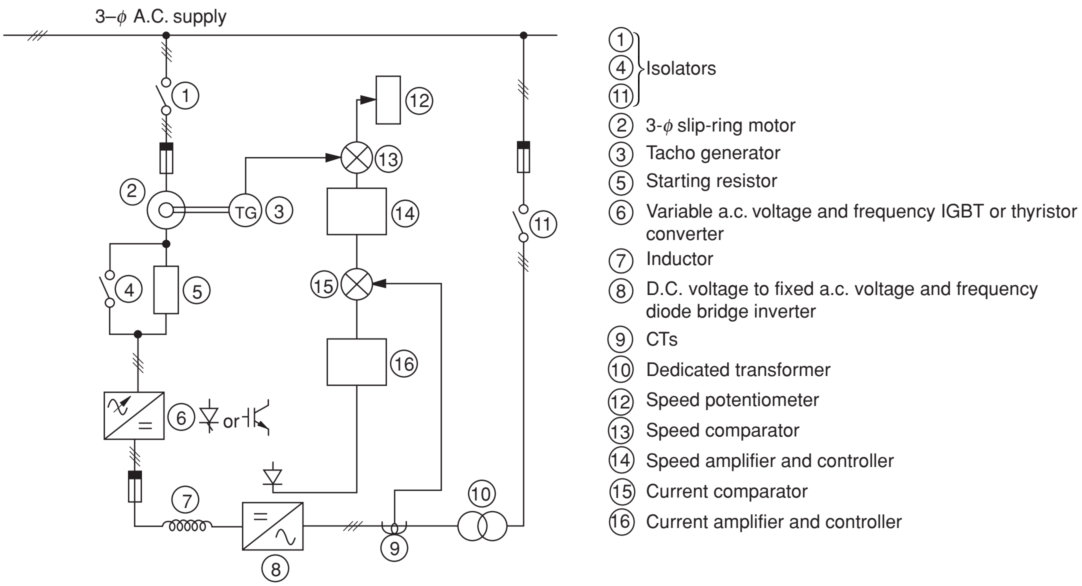



**Figure 6.48** Typical block diagram of a large slip recovery system using IGBTs or thyristors

illustrate a typical slip recovery system and its control scheme, respectively.

The major difference in this configuration from that of a *V/f* control is the variable voltage and frequency from the rotor circuit that is first converted to a d.c. voltage and then inverted to a fixed frequency supply voltage in order to feed the slip power back to the supply source. The converter–inverter combination acts like a variable current source and in turn like a variable resistance. The power saving by this method is twofold. First, the power loss in the external resistance is totally eliminated and second, the rotor power is fed back to the main supply source. This system has a very high initial cost and is therefore preferred for large wound motors above 250 kW. Nevertheless, it is advisable to employ a slip recovery system even for lower rating motors which are required to perform frequent speed variations. Also the regular power saving would offset the heavy initial cost in the first few years only and then provide a recurring energy and cost saving.

## 6.16.4 Operation of a process plant

With the use of static drives for speed control of induction motors, through open or closed-loop feedback control systems, it is now possible to monitor and control a process line automatically, which would not be feasible if carried out manually. We will consider a simple process line of a continuous galvanizing plant to demonstrate the application of this technology in automatic and accurate control of a process industry.

The total engineering of such a system is custom-built. It first requires a thorough study of the process, dividing the process into various activities and then monitoring and controlling each activity through these controls to achieve the required process operation.

Figure 6.49 illustrates a continuous hot-dip galvanizing line to perform zinc coating of MS sheets so that the production line has no discontinuity even when the supply of sheet is exhausted, or during a changeover from one feeding route to another or at the finishing line during a changeover from a completed roll to an empty one. All this is possible with the use of this technology as described below.

The process line indicates only the vital areas. There may be many more auxiliary drives and controls, inter-connected within the same process line, to adjust the process and its quality more closely. They have not been shown in the figure for the sake of simplicity. We do not discuss the duration of one cycle, its speed, the temperature of the furnace or the hot dip zinc vessel etc. and other important parameters. These all are a matter of detailed engineering and process requirements. We describe the process only broadly, to give an idea of applying the technology to a process line for very precise control. We employ encoders to give a pulse output of speed of a particular drive, and PLCs* to implement the process logics through the various drives.

*PLC – Programmable logic controller (the registered trademark of Allen Bradley Co. Inc., USA).

The use of PLCs is essential in the control of motors to closely monitor the operating parameters of the process line on which the motors are connected. The inverter unit controlling the motors then conducts the required correction in each motor speed through its switching logistics, which may be activated by the motor side $V$, $f$, $I_m$, $I_a$ or $\beta$ etc., depending upon the inverter logistics being used. All such parameters are predefined for a particular process and are preset.

### *A brief description of the process and the use of static drives*

We have divided the total process line into three sections:

### Uncoiler section

1. Pay-off reel no. 1 feeds the raw MS sheet to the process section via feed pinch rolls nos 1 and 2 which straighten the sheet before it enters the welder.
2. To maintain continuity and achieve an uninterrupted process line a second parallel feed route is provided through a second pay-off reel no. 2.
3. These rolls are driven by motors $M_1$ and $M_2$ whose speed is controlled through the tension of the travelling sheet. The tension of the sheet is adjusted by monitoring the diminishing diameter of the payoff roll and the thickness of the sheet.
4. The pay-off roll is unwound by the tension of the sheet, caused by the speed of the recoiler at the finishing line and the bridles positioned at different locations. The pay-off roll motors therefore operate in a regenerative mode and can feed back the energy thus saved to the source of supply, if desired. This can be done by using a full-wave synchronous inverter, as shown in Figures 6.31 or 6.33.

   However, this is a more expensive arrangement. A more economical method is to use a full-wave, diode bridge converter and an IGBT inverter unit combination as shown in Figure 6.50 in place of an additional thyristor or IGBT feedback circuit. The d.c. link bus is now made a common bus for all the drives operating on the process. During a regenerative mode, such as during uncoiling the pay-off reels, the voltage of the d.c. bus will rise and will be utilized to feed the other drives. This process will draw less power from the source. The regenerative energy is now utilized in feeding the process system itself rather than feeding back to the source of supply. There is now only one converter of a higher rating, reducing the cost of all converters for individual drives and conserving regenerative energy again at a much lower cost.

   There is no need to introduce a resistance for the purpose of dynamic braking for the individual drives, but a large resistance will be necessary on the d.c. bus to absorb the heat energy during shutdown (braking) of the process. The useful energy during shutdown cannot be fed back to the source due to the configuration of the converter–inverter combination. This arrangement can feed the regenerative energy to its own process only.

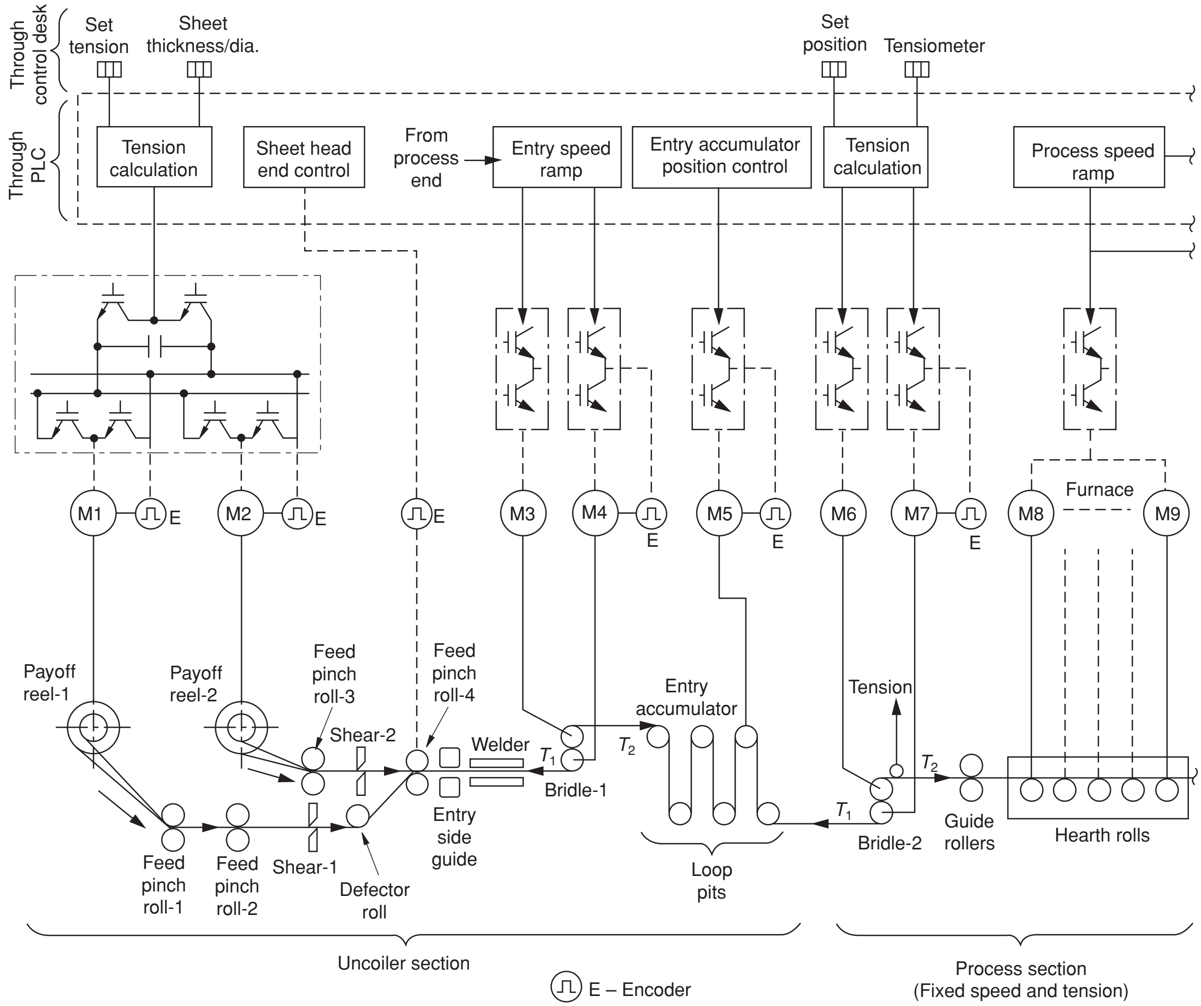



**Figure 6.49** *(Contd.)*

The scheme also facilitates conservation of regenerative energy if there are more of such drives without any additional cost.

5 Shear no. 1 is used to shear the edges of the sheet of pay-off reel no. 1 at the beginning as well as at the end of the coil before it enters the welder to smooth this for a correct welding with the outgoing edge. At the beginning of the coil this edge is welded with the tail end of the previous coil and at the end it is welded with the edge of the fresh coil at the beginning from pay-off reel no. 2. Pay-off reel no. 2, driven by motor $M_2$, is arranged parallel to pay-off reel no. 1 to provide a second feed route for an uninterrupted and continuous process flow.

6 The deflector roll guides the sheet to another pinch roll no. 4 to carry out precise alignment of the edges before they enter the welder.

7 Pinch roll no. 4 aligns the edge of the sheet through a feedback control.

8 For further alignment of the edge width just before entering the welder the sheet is guided again through an entry-side guide.

9 With the help of bridle no. 1 driven by motors $M_3$ and $M_4$, the uncoiler section speed is controlled by monitoring the tension of the travelling sheet and hence maintaining constant speed of the sheet in the uncoiler section. The tensile difference of $T_1$ and $T_2$ determines the speed of the uncoiler. Speed and tension of the sheet must remain constant for absolute synchronization between the uncoiler process and the recoiler sections.

10 To allow for welding time, a buffer of a certain sheet length, in the form of entry accumulator, is maintained, generally in a vertical formation, to save space. This feeds the line ahead until the welding operation is completed and the second route is installed to feed the process.

11 The time gap in carrying out the welding is

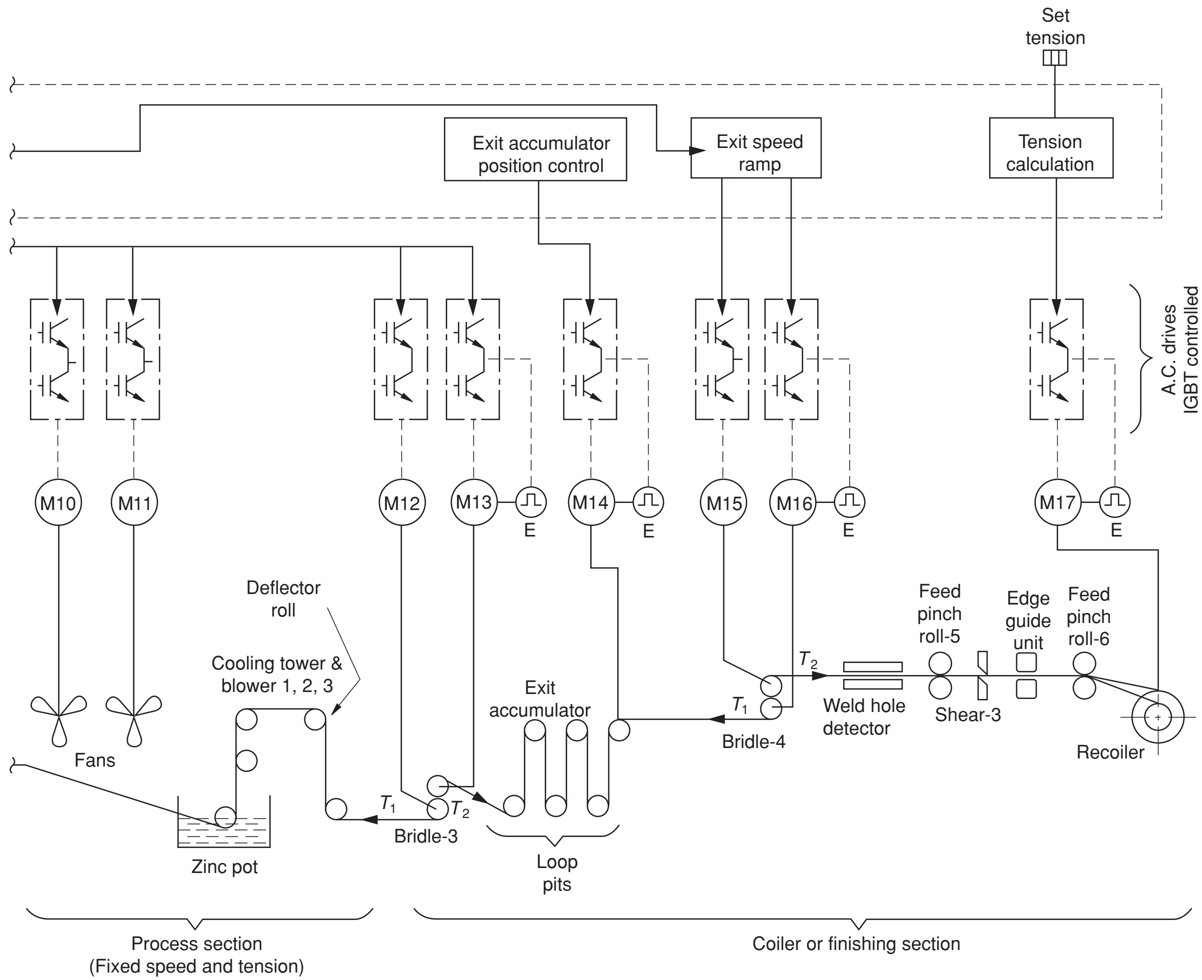



**Figure 6.49** Typical process layout of a continuous hot dip galvanizing line

compensated by raising the speed of the second route now introduced until the predefined buffer of an excess length of sheet is produced with the help of accumulator drive motor $M_5$.

## Process section

12 The tension and speed in the process section is maintained again with the help of bridle no. 2, driven by motors $M_6$ and $M_7$.

13 The sheet is now fed through a pair of guide rollers to a furnace section through a de-greasing tank, where it is preheated for drying and raising the temperature of the sheet up to a required level (400–465°C typical) before it enters the hot galvanizing pot for the desired thickness of zinc coating. The movement of the sheet through the furnace is helped by motors $M_8$ and $M_9$.

14 The hot-treated sheet is cooled to the required level by fans, driven by motors $M_{10}$ and $M_{11}$ before it enters the molten zinc pot.

15 The required thickness of zinc coating is achieved by dipping it in the molten zinc pot for a preset time. The thickness of the coating is monitored by two rollers through which the coated sheet is passed. The time of welding, degreasing, heating and hot dipping are synchronized so that the welding operation takes the same time as the degreasing, heating and hot dipping.

16 Bridle no. 3, driven by motors $M_{12}$ and $M_{13}$, is the process section master controller and controls the speed and travel of the sheet in the process section.

## Coiler or the finishing section

This section is almost the same as the uncoiler section.

17 The hot galvanized sheet is cooled by blowers and fed to an exit accumulator driven by motor $M_{14}$, similar to the entry accumulator, via a deflector roll to adjust its position.

18 Before the finished sheet is finally cut into lengths as required or rolled into recoilers its exit speed and tension is monitored and controlled again by bridle no. 4, driven by motors $M_{15}$ and $M_{16}$.

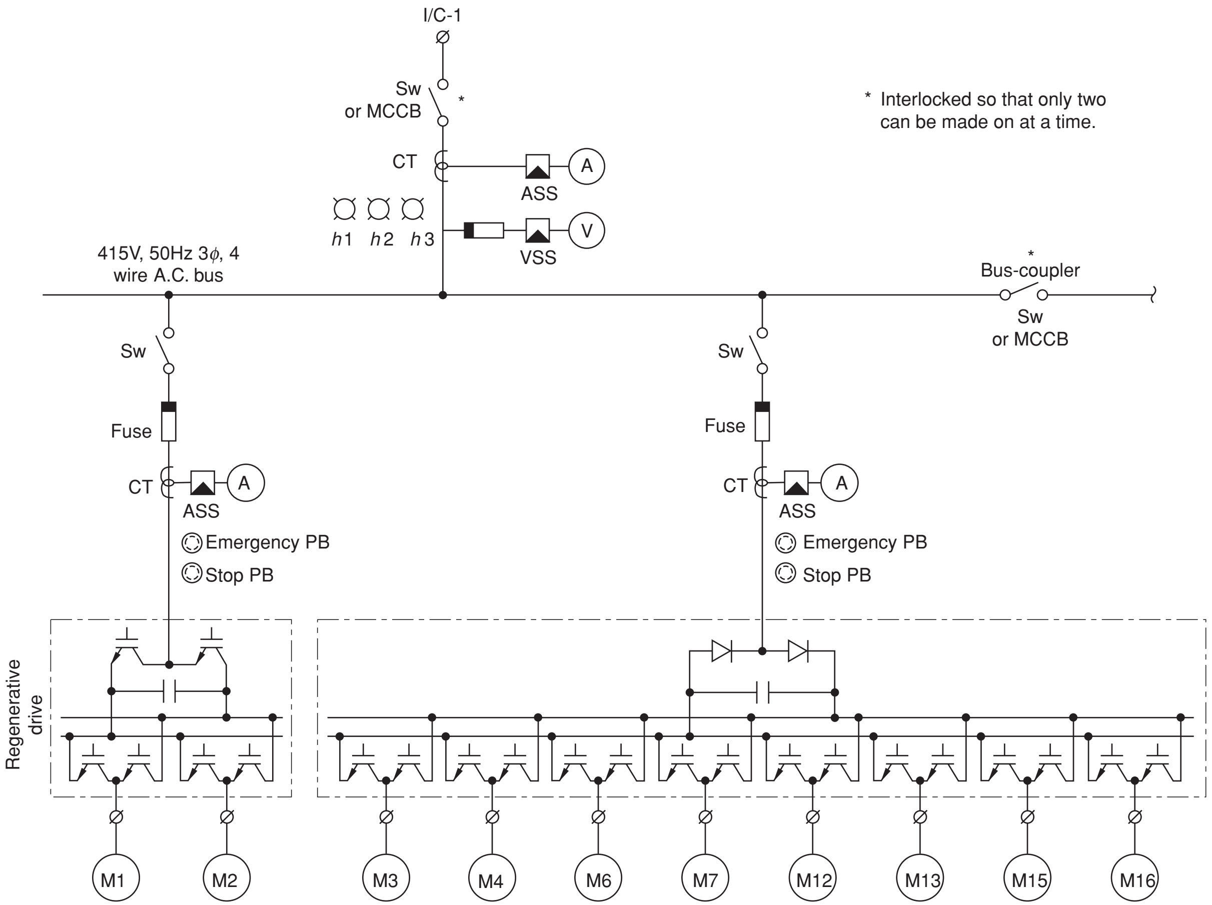



**Figure 6.50** *(Contd.)*

19 The sheet may also be inspected for the quality of welding and checked for pin-holes by a weld-hole detector before it is finally cut into lengths or wound onto rolls.
20 It then passes through a pair of guide rollers, pinch rollers no. 5, a shear, an edge guide unit (to align the width of the sheet) and a final alignment pinch roll no. 6 as shown.
21 The recoiler is driven by motor $M_{17}$ that adjusts its speed and tension as calculated for the whole process line. It is this drive and the bridles that maintain the required tension throughout the process and make the pay-off reel drives operate as regenerative units.

*Note*

- We have considered all drives as a.c. although d.c. drives are also in use. Earlier only d.c. drives were used.
- IGBT inverters have been considered.
- All activities can be monitored through a control desk.
- All controls and precise adjustments are carried out by a PLC.

### 6.16.5 Other applications

In view of the low maintenance cost and high reliability of solid-state devices, combined with precision and accuracy through a feedback network, this technology finds many applications in such process industries as the following:

- Textiles
- Sheet and plate metal working
- Packaging
- Material handling
- Steel rolling
- On- and off-shore drilling platforms
- Machine tools
- Process industries such as sugar, printing machinery, cement mills, chemicals, paper
- Thermal power plant auxiliaries such as flow control of primary air fan, ID fan and forced-draught fans, boiler feed pumps, circulating water pumps and condensate pumps, coal handling plant (e.g. ball mill, wagon tippler, and stacker reclaimer)
- Mining. A solid-state semiconductor device has no physical contacts to make or break the current. There is thus no arc formation during switching of these devices. Therefore they have a very wide application in mines and other hazardous areas using flameproof a.c. motors where d.c. machines cannot be used.

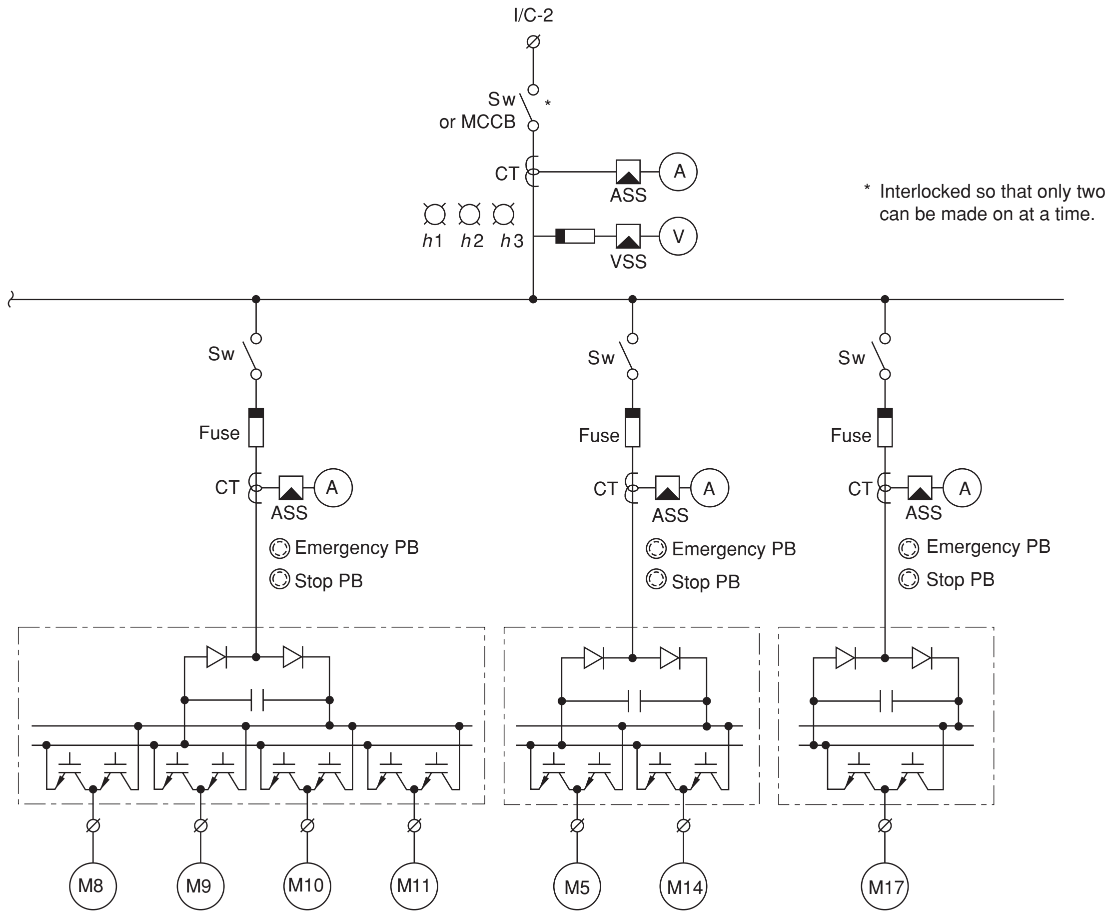



**Figure 6.50** Power distribution arrangement for the galvanizing process of Figure 6.49

## 6.17 Speed variation through variable-speed fluid couplings

The speed of the motor can also be varied by a variable-speed fluid drive as discussed in Section 8.4.

## 6.18 Static drive versus fluid coupling

Variable-speed drives are essential for many industrial applications requiring variable operating parameters during the course of operation. Such variations can be in the flow of fluid and pressure of air or gas etc. The conventional method of throttle control through a vane or a damper causes a considerable waste of energy. To obtain a variable speed and yet save on energy, one can use either a static drive as discussed earlier or a variable speed fluid coupling (Chapter 8). The choice between the two will be a matter of system requirements and an overall assessment of the ease of application, economy and accuracy of speed control in addition to the amount of energy it will be able to conserve. The application engineer would be a better judge to make a more appropriate choice between the two, based on system requirements. We give a brief comparison between these drives in Table 6.5. This comparison should help in making a more judicious choice of drive.

Fluid couplings larger than, say, 400 kW can be employed with ease in achieving the required speed control. They also provide a mechanical coupling between the drive and the load. A variable-speed fluid coupling also results in energy saving along the lines as discussed in Example 6.1. In many countries this device is classified as ‘energy saving’ and attracts subsidies from the state. They are simple to operate and are highly economical in their initial cost compared to static drives. The only case against their use, despite the exorbitant costs of static drives, is their recurring power losses, as mentioned in Table 6.5 (items 19 and 20). It is a constant and recurring drain on useful energy. A static drive, irrespective of its cost, has a pay-off period of three to four years, depending upon the size of drive, frequency and duration of speed control and its accuracy. Thereafter it achieves regular and high energy saving. This advantage is not possible

**Table 6.5** Comparative study of performance of an a.c. drive, variable speed fluid coupling and a d.c. drive

| *Features* | *A.C. drive* | *Variable speed fluid coupling* | *D.C. drive* |
|---|---|---|---|
| 1 Manufacturing range | Any range | 40 to 30 000 kW by USA | 0.1 to 10 000 kW |
| 2 Starting acceleration | Soft and stepless start | Soft and stepless start | Soft and stepless start |
| 3 Starting current ($I_{st}$) | Can be controlled to any desired level subject to the minimum $T_{st}$ required | Cannot be controlled. Although duration of the starting inrush is very short because of no-load start | Can be controlled to any desired level. Normally up to 1.5 times the rated current |
| 4 Starting torque ($T_{st}$) | As desired | As desired | As desired up to 5–6 times the $T_r$, through field forcing (i.e. by raising the field voltage during the start up period) |
| 5 Variation in torque with speed | • Through *V/f* control, the torque can be kept constant $\left(T \propto \phi_m \cdot I_r \propto \frac{e_2}{f_r} = \text{constant}\right)$ up to a desired speed, *N*. Since h.p. ∝ T.N., h.p. varies with speed<br>• Beyond *N*, the h.p. can be kept constant by keeping the voltage fixed and raising *f* (h.p. ∝ T · N). The speed-torque characteristic is similar to a d.c. machine as shown in Figure 6.51 | $T \propto N^2$ | Characteristics are similar to an induction machine as shown in Figure 6.51:<br>(i) Up to the base speed *N*, torque can be kept constant through armature voltage control, $V \propto I_a$, field current ($I_m \propto \phi$), remaining constant ($T \propto \frac{\phi \cdot I_a}{N}$ = constant, $I_a$, being the armature current)<br>(ii) Beyond the base speed *N*, h.p. can be kept constant (h.p. ∝ T · N) by reducing the field current ($I_m \propto \phi$) and keeping the armature voltage as constant ($I_a$ = constant). Torque will now diminish exponentially, since $\phi$ diminishes and *N* rises, *N* being in denominator, has the same effect as $\phi$ (Figure 6.51) |
| 6 Magnetising losses of the motor | Vary with *f* | Remain at 100% | Remain constant for speed variations within the required speed *N*, as the field current is kept fixed and only the armature voltage is varied. For speed variations beyond *N*, however, when the armature voltage is kept constant and the field current is varied, the magnetizing losses also vary |
| 7 Copper ($I^2R$) losses | Low at reduced speeds, due to low magnetizing current ($I_m$) and correspondingly lower $I_1$ | No reduction because of same magnetizing losses and therefore relatively higher $I_1$ | At lower speeds more than the a.c. drives as $I_m$ remains the same and a relatively higher $I_a$ |
| 8 Power factor | Although $I_m$ is low, overall p.f. may be slightly lower on the line side, because of harmonic contents and inclusion of *L*. *L* is introduced to (i) Limit the current harmonics and (ii) Limit the rate of rise of current, i.e. ripples (d*i*/d*t*) | Low as $I_m$ remains the same | Lower than the a.c. drive because of same field current |
| 9 Combined efficiency of the motor and the drive | At rated speed 90% and more. Reduces slightly at lower speeds because of poor efficiency of the | (i) At rated speed ≃ 87–90% (coupling efficiency as high as 97–98%) | At rated speed up to 80–90%. At lower speeds reduces more than the a.c. drives because of |

*(Contd.)*

| | *Features* | *A.C. drive* | *Variable speed fluid coupling* | *D.C. drive* |
|---|---|---|---|---|
| | | machine at lower speeds. Losses in the a.c. controls do not normally exceed 0.5–1.5% | (ii) At two thirds of the input speed $\simeq$ 50%<br>(iii) at 20% of the input speed $\simeq$ 66%. See Figure 6.52 | fixed field losses |
| 10 | Voltage dip during start-up | Nil | High because of same $I_{st}$ but for a very short duration as the motor picks up lightly | Nil |
| 11 | Fault level | Low | High because of high $I_{st}$ | Low |
| 12 | Any cost reduction in electricals (motor, cables, switchgears etc.). | Yes; because of lower capacity of motor, cables and switchgears and a low fault level | Similar cost reduction possible but all requirements to be suitable for slightly higher fault level because of high $I_{st}$ | Yes, as in a.c. drives |
| 13 | Range of speed control | Very wide and stepless up to zero speed | Moderate to accurate, depending upon the accuracy of controls. Stepless up to 20% of $N_r$ at constant h.p. and up to 33% of $N_r$ at constant torque is possible. Pumps, ID fans etc., that call for speed variation during a process need may not necessarily be too accurate. Or variation in flow of fluid, gas or temperature etc. not calling for very accurate controls, that such drives find their extensive use. It may be made more accurate, but at higher cost of controls | Very wide and stepless as for a.c. drives |
| 14 | Accuracy of speed control | Up to ± 0.01% in open-loop and ± 0.001% in closed-loop control systems | Moderate to precise controls as for a.c. drives possible with the use of microprocessor-based control systems | Very accurate speed controls up to ±0.01% |
| 15 | Monitoring of operating parameters | Very accurate controls through microprocessor-based closed-loop feedback control systems | Moderate to microprocessor-based, fully programmable logic controls and feedback control systems are available, to provide smooth speed controls as good as a.c. drives | Same as for a.c. drives |
| 16 | Acceleration and braking | Possible | Possible | Possible |
| 17 | Reversal | Possible | Not possible. Although coupling is bidirectional, it can be run in any one direction only | Possible |
| 18 | Inching | Possible | Possible | Possible |
| 19 | Power loss | No loss except in the form of motor inefficiency at lower speeds | Relatively higher losses because of<br>(i) High starting current<br>(ii) Coupling slip up to 15–16% at two thirds the input speed and about 20% at 20% of the input speed (slip reduces at lower speeds as illustrated in Figure 6.52 | Losses are high because of field system, but comparatively much less than fluid and eddy current couplings. At lower speeds the losses rise in the form of motor inefficiency |
| 20 | Energy saving | Optimum saving up to 100% (no loss) | Good saving. But low compared to a.c. drives because of high slip losses. As it saves energy, it is also entitled to state subsidies | Slightly low, 90–94%, because of field losses (up to 5–7%) |

*(Contd.)*

| *Features* | *A.C. drive* | *Variable speed fluid coupling* | *D.C. drive* |
|---|---|---|---|
| 21 Harmonics | They generate high harmonics $nk \pm 1$ (Section 23.6(b)) and must be suppressed by using filter circuits | There arises no such phenomenon | Higher than even a.c. drives, because of commutator arcing which it produces at high frequencies |
| 22 Unfriendly environmental conditions such as dustladen areas, fire hazardous, corrosive and contaminated locations | Not suitable. The static controls must be located away from such areas in well-protected rooms | No problems as it is a sealed unit. But for controls which may be located separately | Not suitable at all, as the motor itself cannot be relocated in safer areas |
| 23 Heating | (i) Moderate in transistor controls but<br>(ii) Excessive in thyristor controls. The higher the rating the higher the temperature rise. They require extensive cooling arrangements which may be external or forced | Very high at lower speeds and requires extensive cooling | Moderate |
| 24 Maintenance and downtime | May be high. For efficient maintenance high-skilled operators and their proper training are essential, besides stocking enough spares as recommended by the manufacturer. Yet the expert services of the manufacturer may sometimes be necessary. At times, the spares may not be readily available or it may not be possible to repair them immediately. In such cases either it is a total shutdown or the controls have to be bypassed and the machine run on DOL<br>*Note:* With the availability of remote service facility with the drive manufacturers through remote drive and telemetry software as noted in Section (6.7.4(5)), it is now easy to communicate with the manufacturer and avail of their instant on line expert services in most cases. | Very low maintenance. Where a mechanical ruggedness is required, a fluid coupling provides a more reliable solution | Downtime due to brushes is about 1 hour only in 6 months' to a year's time. However, meticulous monitoring of brushes and commutator condition is important and so also the environmental conditions. Controls will also need similar attention as for a.c. drives. In a breakdown, the motor cannot be run in any way (unlike a.c. motors) and would lead to a total shutdown |
| 25 Cost consideration | A costly arrangement | It is generally cheaper than a static drive in all ranges | Quite economical compared to an a.c. drive. The following may give a rough idea:<br>Up to 10 kW – price difference not significant<br>Above 10–40 kW– a.c controls may be costlier by 30–40%<br>Above 40–130 kW – a.c. controls may be costlier by 40–50%<br>In the range of 500 kW and above – a.c. controls may be costlier by 2.5 to 3 times and even more in still higher ranges. Remote service facility is available here also. |

in a fluid coupling due to higher power losses. Conservation of energy is therefore the main criterion in the selection of static drives.

Until a few years ago, when static technology was still in its infancy, variable-speed fluid drives had very wide application. With the advent of static technology, the trend is shifting in favour of static controls, particularly for drives that have to undergo frequent and wide speed variations during their normal course of operation, or which require very accurate speed control. For applications not needing very precise speed control or wide variations in speed (e.g. high-capacity pumps or ID fans) variable-speed fluid couplings are still the best choice.

## 6.19 D.C. drives

The use of d.c. motors for precise speed controls is long practised and it had been a unanimous choice until a few years ago. It still is, in a few applications, purely on cost consideration. But its use is now gradually waning out due to their maintenance and down-time which a process industry can ill afford and also more advanced technology now being available in static controls to control an a.c. motor. An induction motor is maintenance free. But older installations still use d.c. motors and may continue to do so for a few more years, until they go in for a retrofitting of the existing d.c. motor system with an a.c. motor and drive system or the next modernization of the installation. Many leading d.c. motor manufacturers have already discontinued the manufacture of such machines due to a sharp decline in demand. There are a few manufacturers still in the field and may continue until would exist a demand for replacements and extensions of existing load lines. A few manufacturers who have discontinued the production of this machine have established links with those still in the field to cater for replacements. Therefore this book has dealt only briefly with this machine. Figure 6.32 shows a converter circuit for the control of a d.c. machine, also showing a fully controlled thyristor inverter unit for regenerative energy feedback. Figure 6.50(a) shows the outside view of a digital converter d.c. drive.


**Figure 6.50(a)** Digital converter DC drive (Courtesy: BCH)

To make a better comparison between an a.c. and a d.c. drive we illustrate in Figure 6.51 for a d.c. motor, the likely variation in its torque, with variation in the applied voltage, below the base speed and with a constant voltage but variable field strength, above the base speed.

## 6.20 Retrofitting of EE motors and drives

Substantial energy and consequent generation of heat can be saved (recurring cost saving apart) by adapting to EE motors and also employing variable speed drives when motors are to perform variable duties. The payback period, irrespective of cost of retrofitting may not exceed

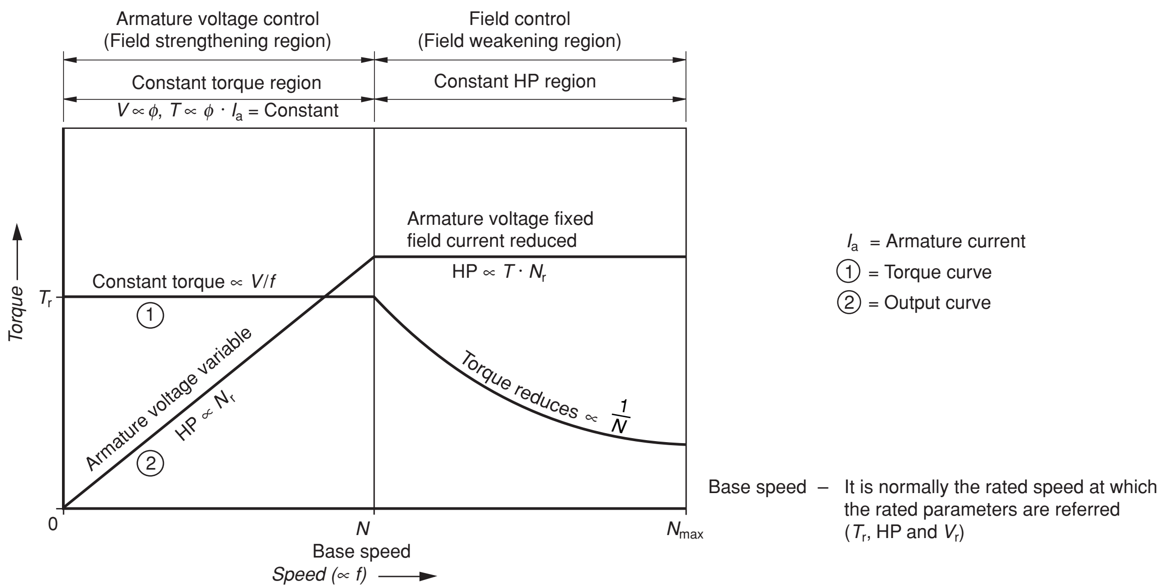



**Figure 6.51** Variation of torque with speed in a d.c. machine (the same as for an a.c. machine)

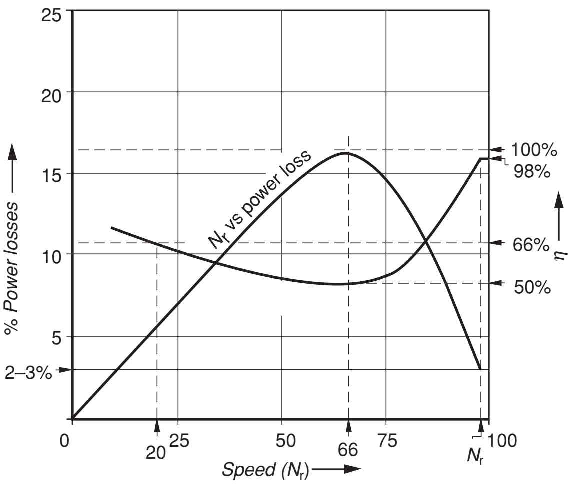



**Figure 6.52** An approximate illustration of $\eta$ and loss variation, with change in speed in a variable-speed fluid coupling

a year or so depending upon the running hours/year. If the motor was running under-loaded before the retrofitting, the payback period may even be less because most energy saved now, was being consumed and paid for before. It is therefore recommended to replace the existing motors with EE motors and conventional starting to soft staring when it is a fixed load drive or a variable speed drive when the load is fluctuating. It is strongly recommended that motor and switching device both be retrofitted for optimum advantage. It may even be done in phases – in first changing the motor to EE motor suitable for inverter duty and in the second retrofitting the drive (taking care of cable length). Providing only a drive without changing the motor is a job half done because,

– The old motor even if it is not oversized is still a drain on useful energy by way of higher losses and lower efficiency even if it operates under controlled supply parameters just adequate to meet its duty requirements.
– The insulation level of the motor may have aged out and may not be adequate to meet the PWM switching surges on the motor side.
– The inverter drives cause additional heating to the motor windings due to high harmonic contents and may call for a derating of the motor. It is possible that the motor was chosen over-sized initially but has no reserve capacity now as the load demand may have risen over the period. If it is still adequate to meet the present load demand, it is possible that it may not be suitable for inverter duty causing extra heating. In most cases, however, this may not be necessary as the motors usually have enough reserve capacity. In any case analysis for motor suitability is mandatory.
– The old motors would call for bearing insulation whenever bearing parasitic currents may exist as discussed before.

Therefore, it is advisable that with drives, existing motors may also be replaced with EE motors suitable for inverter duty. In certain applications close co-ordination between the motor, the drive and the length of the inter-connecting cable may be mandatory to ensure total compatibility and safety to perform the desired duty within safe parameters and also ensure safety at hazardous locations. See also Section 7.17.1.

**Relevant Standards**

| *IEC* | *Title* | *IS* | *BS* | *ISO* |
|---|---|---|---|---|
| 60034-1/2004 | Rotating electrical machines. Rating and performance. | 4722/2001,325/2002 | BS EN 60034-1/1998 | – |
| 60034-17/2003 | Cage induction motors when fed from converters. Application guide. | – | – | – |
| 60146-1-1/1991 | Specifications of basic requirements for power converters. | 14256/2000 | BS EN 60146-1-1/1993 | – |
| 60439-1/2004 | Low voltage switchgear and controlgear assemblies. Requirements for type-tested and partially type tested assemblies. | 8623-1/1998 | BS EN 60439-1/1999 | – |
| 60947-4-2/2002 | Low voltage switchgear and controlgear – contactors and motor starters – a.c. semiconductor motor contactors and starters. | – | – | – |
| 60947-4-3/2000 | Low voltage switchgear and controlgear – contactors and motor starters – a.c. semiconductor controllers and contactors for non-motor loads. | – | – | – |
| 61000-5-2/1997 | Electromagnetic compatibility (EMC) Part 5 – Installation and mitigation guidelines Section 2 – Earthing and cabling. | – | – | – |
| — | Code of practice for selection, installation and maintenance of switchgear and controlgear: Part 3 – Installation. | 10118-2,3/2001 | – | – |
| 61800-3/2004 | Adjustable speed electrical power drive systems. EMC requirements and specific test methods. | – | BS EN 61800-3/2004 | – |

**Related US Standards ANSI/NEMA and IEEE**

| | |
|---|---|
| ANSI/IEEE-142/1991 | Grounding of industrial and commercial power systems. (*IEEE Green Book*) |
| ANSI/IEEE-519/1993 | Guide for harmonic control and reactive compensation of static power converters. |
| ANSI/IEEE-C37.21/1998 | Standard for control switchboards. |
| ANSI/IEEE-C62.41/1991 | Recommended practice for surge voltages in low voltage a.c. power circuits. |
| NEMA/ICS-1.1/1998 | Safety guidelines for the application, installation and maintenance of solid state control. |
| NEMA/ICS-7/2000 | Industrial control and systems. Adjustable speed drives. |
| NEMA/ICS-7.1/2001 | Safety standards for construction and guide for selection, installation and operation of adjustable speed drive systems. |
| NEMA/ICS-9/1993 | Industrial control and systems. Power circuit accessories (requirement for brakes). |
| NEMA/ICS-16/2001 | Motion/position control motors and cables. |
| NEMA/LS-1/1992 | Low voltage surge protection devices. |
| NEMA/MG-1,Part 30/1998 | Application considerations for constant speed motors used on a sinusoidal bus with harmonic contents |
| NEMA/MG-1,Part 31/1998 | Motors and generators ‘Definite-purpose inverter-fed polyphase motors’ |
| DIN 6-1800-3/2002 | Cage induction motors when fed from converters. Application guide. |

*Notes*

1 In the table of relevant Standards while the latest editions of the Standards are provided, it is possible that revised editions have become available or some of them are even withdrawn. With the advances in technology and/or its application, the upgrading of Standards is a continuous process by different Standards organizations. It is therefore advisable that for more authentic references, one may consult the relevant organizations for the latest version of a Standard.

2 Some of the BS or IS Standards mentioned against IEC may not be identical.

3 The year noted against each Standard may also refer to the year it was last reaffirmed and not necessarily the year of publication.

## List of formulae used

### Speed control through phasor control

$$\bar{I}_1 = \bar{I}'_m + \bar{I}_m + \bar{I}_a \tag{6.1}$$

$I_1$ = line current
$I'_m$ = loss component
$I_m$ = magnetizing component
$I_a$ = active component

### Field-oriented control

$$T = k \cdot I_m \cdot I_a \sin\theta \tag{6.2}$$

$\theta$ = phasor displacement between $\bar{I}_m$ and $\bar{I}_a$ (electrical position of rotor field in space with respect to stator)

### To obtain variable *V* and *f* in IGBTs through PWM

$$CDF = \frac{t_1 + t_2 + t_3 + t_4 + t_5 + t_6}{T} \tag{6.3}$$

$t_1$, $t_2$ … $t_6$ are the pulse widths in one half cycle
$T$ = one half of a cycle

$$V_{r.m.s} = V\sqrt{CDF} \tag{6.4}$$

$V$ = amplitude of output, voltage pulses
$V_{r.m.s}$ = r.m.s. value of the output, a.c. voltage

### To smooth output a.c. ripples

$$Q = C\frac{dv}{dt} \tag{6.5}$$

$Q$ = charge stored by the capacitor unit
$C$ = capacitance of the capacitor
$dv/dt$ = rate of voltage change or a.c. ripples in the d.c. link

### Current source inverter (CSI) to vary $I_1$ and $f$

$$V \simeq L\frac{di}{dt} \text{ (ignoring } R \text{ of the circuit)} \tag{6.6}$$

$V$ = voltage across the inductor
$L$ = large series inductor
$\frac{di}{dt}$ = a.c. ripples

### Computation of energy saving in a pump

$$P = \frac{H_d \cdot Q \cdot d}{36 \cdot \eta} \tag{6.7}$$

$P$ = shaft input in kW
$H_d$ = head in bar
$Q$ = discharge in $m^3$/hour
$d$ = specific gravity of the liquid in $g/cm^3$
$\eta$ = efficiency of the pump

## Further Reading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