Learning Outcome 1.1

State the necessary conditions to produce a rotating magnetic field from stationary coils energised with an AC Supply

10.11 Rotating Field due to a Three-phase Winding

Fig. 10.17 shows a stator winding with three diametral coils aa', bb' and cc', each having N_s turns. The dots and crosses indicate the direction of conventionally positive current in each coil as explained

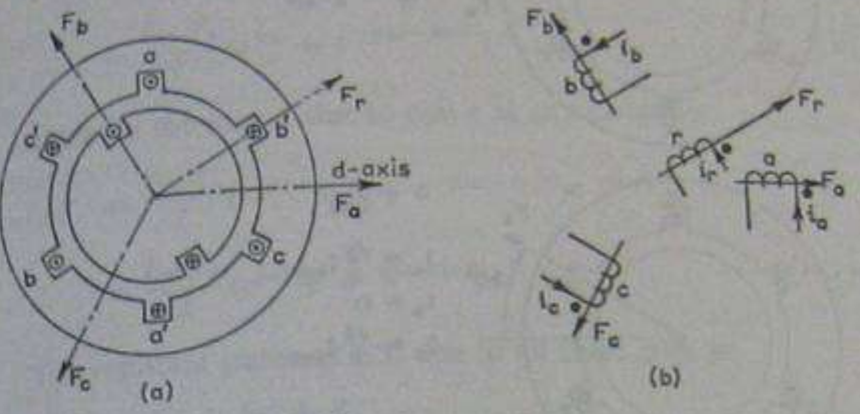


Fig. 10.17 M.M.F. DUE TO A 3-PHASE WINDING

in Section 10.3. The axes of the coil m.m.f.s are therefore mutually displaced by $2\pi/3$ radians, as shown in Fig. 10.17.

Suppose the three coils are supplied with balanced 3-phase currents, i_a, i_b and i_c, such that

$$i_a = I_{sm} \cos \omega t = \frac{I_{sm}}{2} (e^{j\omega t} + e^{-j\omega t}) \tag{10.44}$$

$$i_b = I_{sm} \cos (\omega t - 2\pi/3) = \frac{I_{sm}}{2} (e^{j(\omega t - 2\pi/3)} + e^{-j(\omega t - 2\pi/3)}) \tag{10.45}$$

$$i_c = I_{sm} \cos (\omega t + 2\pi/3) = \frac{I_{sm}}{2} (e^{j(\omega t + 2\pi/3)} + e^{-j(\omega t + 2\pi/3)}) \tag{10.46}$$

The m.m.f. of coil a is directed in the reference direction when i_a is positive. The instantaneous value of this m.m.f. is therefore

$$F_a' = \frac{I_{sm}N_s}{2} (e^{j\omega t} + e^{-j\omega t})e^{j0} \qquad (10.47)^*$$

This expression has been multiplied by e^{j0} $(= 1)$ to indicate that it acts in the space reference direction.

* To avoid confusion with f for frequency, instantaneous m.m.f. will be represented by F'.

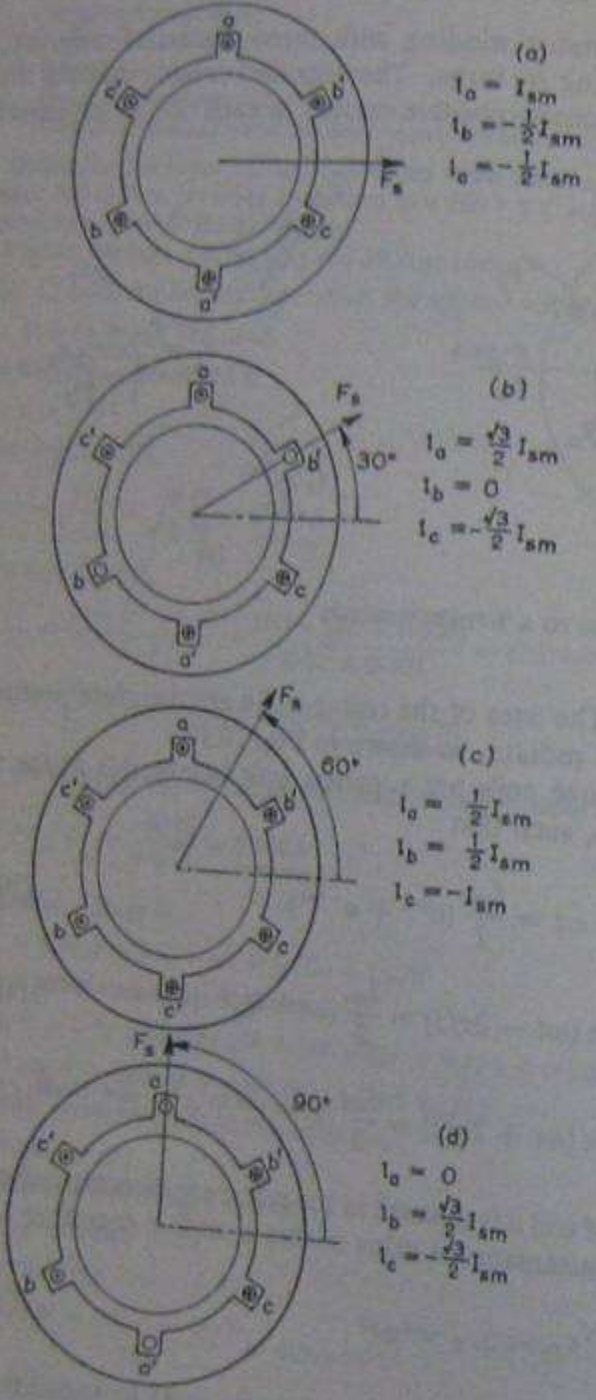


Fig. 10.18 M.M.F. DUE TO A 3-PHASE WINDING AT DIFFERENT INSTANTS

The m.m.f. of coil b is directed along an axis $+2\pi/3$ radians from the reference direction when i_b is positive. The instantaneous value of this m.m.f. is therefore

$$F_b' = \frac{I_{sm}N_s}{2}\,(e^{j(\omega t-2\pi/3)} + e^{-j(\omega t-2\pi/3)})e^{j2\pi/3}$$

$$= \frac{I_{sm}N_s}{2}\,(e^{j\omega t} + e^{-j(\omega t-4\pi/3)}) \qquad (10.48)$$

Similarly the m.m.f. due to coil c at any instant is

$$F_c' = \frac{I_{sm}N_s}{2}\,(e^{j(\omega t+2\pi/3)} + e^{-j(\omega t+2\pi/3)})e^{-j2\pi/3}$$

$$= \frac{I_{sm}N_s}{2}\,(e^{j\omega t} + e^{-j(\omega t+4\pi/3)}) \qquad (10.49)$$

The resultant stator m.m.f. due to all three coils is

$$F_s' = F_a' + F_b' + F_c'$$

$$= \frac{I_{sm}N_s}{2}\,[e^{j\omega t} + e^{-j\omega t} + e^{j\omega t} + e^{-j(\omega t-4\pi/3)} + e^{j\omega t} + e^{-j(\omega t+4\pi/3)}]$$

Since $e^{-j\omega t} + e^{-j(\omega t-4\pi/3)} + e^{-j(\omega t+4\pi/3)} = 0$,

$$F_s' = \tfrac{3}{2}\, I_{sm}N_s e^{j\omega t} \qquad (10.50)$$

This equation shows that, when three coils are so positioned that their m.m.f. axes are mutually displaced by $2\pi/3$ radians and are then supplied with balanced 3-phase currents, an m.m.f. of constant magnitude results and the m.m.f. axis rotates at an angular velocity of ω radians per second.

For the coil configuration and phase sequence chosen the direction of rotation is in the $+\theta$ direction. It will be found that, if the phase sequence is reversed, the direction of rotation of the resultant m.m.f. axis is also reversed.

Fig. 10.18 shows the m.m.f. due to a 3-phase winding supplied with balanced 3-phase currents for a number of different instants. At (a) the current in phase a is positive maximum value and the currents in the two other phases are half the negative maximum value. The negative currents are indicated by showing the current in the cross direction in coil sides b and c, and in the dot direction in coil sides b' and c'. F_s is shown acting along the stator m.m.f. axis.

Figs. 10.18(b), (c) and (d) show successive instants in the 3-phase cycle corresponding to 30° rotations of the complexor diagram.

Learning Outcome 1.2

Calculate the synchronous speed of the rotating magnetic field given the supply frequency and number poles

13.2 Principle of operation

The operation of a 3-phase induction motor is based upon the application of Faraday's Law and the Lorentz force on a conductor (Sections 2.20, 2.21, and 2.22). The behavior can readily be understood by means of the following example.

Consider a series of conductors of length *l*, whose extremities are short-circuited by two bars A and B (Fig. 13.5a). A permanent magnet placed above this conducting ladder, moves rapidly to the right at a speed *v*, so that its magnetic field *B* sweeps across the conductors. The following sequence of events then takes place:

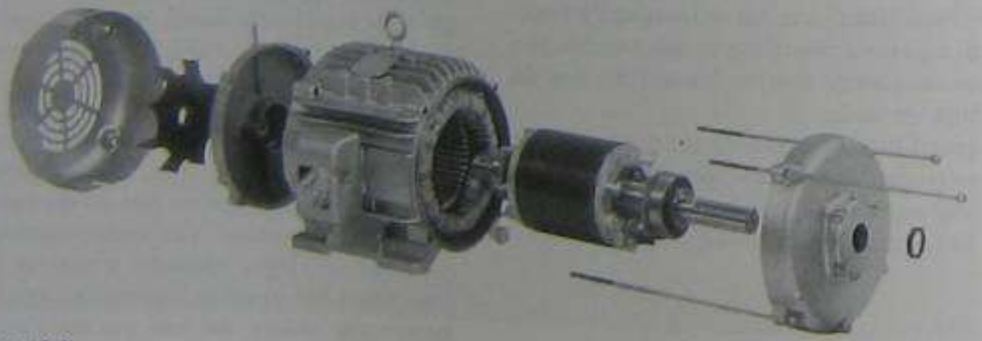


Figure 13.2
Exploded view of the cage motor of Fig. 13.1, showing the stator, rotor, end-bells, cooling fan, ball bearings, and terminal box. The fan blows air over the stator frame, which is ribbed to improve heat transfer. *(Courtesy of Baldor Electric Company)*

Figure 13.3a
Die-cast aluminum squirrel-cage rotor with integral cooling fan.
(Courtesy of Lab-Volt)

1. A voltage $E = Blv$ is induced in each conductor while it is being cut by the flux (Faraday's law).
2. The induced voltage immediately produces a current I, which flows down the conductor underneath the pole-face, through the end-bars, and back through the other conductors.
3. Because the current-carrying conductor lies in the magnetic field of the permanent magnet, it experiences a mechanical force (Lorentz force).
4. The force always acts in a direction to drag the conductor along with the magnetic field (Section 2.23).

If the conducting ladder is free to move, it will accelerate toward the right. However, as it picks up speed, the conductors will be cut less rapidly by the moving magnet, with the result that the induced voltage E and the current I will diminish. Consequently, the force acting on the conductors will also decrease. If the ladder were to move at the same speed as the magnetic field, the induced voltage E, the current I, and the force dragging the ladder along would all become zero.

In an induction motor the ladder is closed upon itself to form a squirrel-cage (Fig. 13.5b) and the moving magnet is replaced by a rotating field. The field is produced by the 3-phase currents that flow in the stator windings, as we will now explain.

13.3 The rotating field

Consider a simple stator having 6 salient poles, each of which carries a coil having 5 turns (Fig. 13.6). Coils that are diametrically opposite are connected in series by means of three jumpers that respectively connect terminals a-a, b-b, and c-c. This creates three identical sets of windings AN, BN, CN, that are mechanically spaced at 120° to each other. The

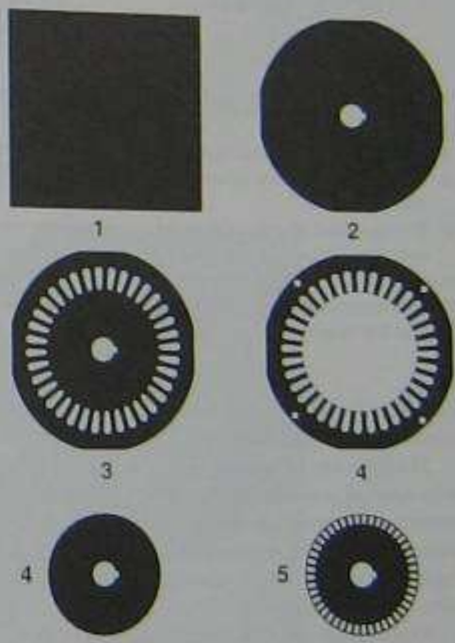


Figure 13.3b
Progressive steps in the manufacture of stator and rotor laminations. Sheet steel is sheared to size (1), blanked (2), punched (3), blanked (4), and punched (5).
(Courtesy of Lab-Volt)

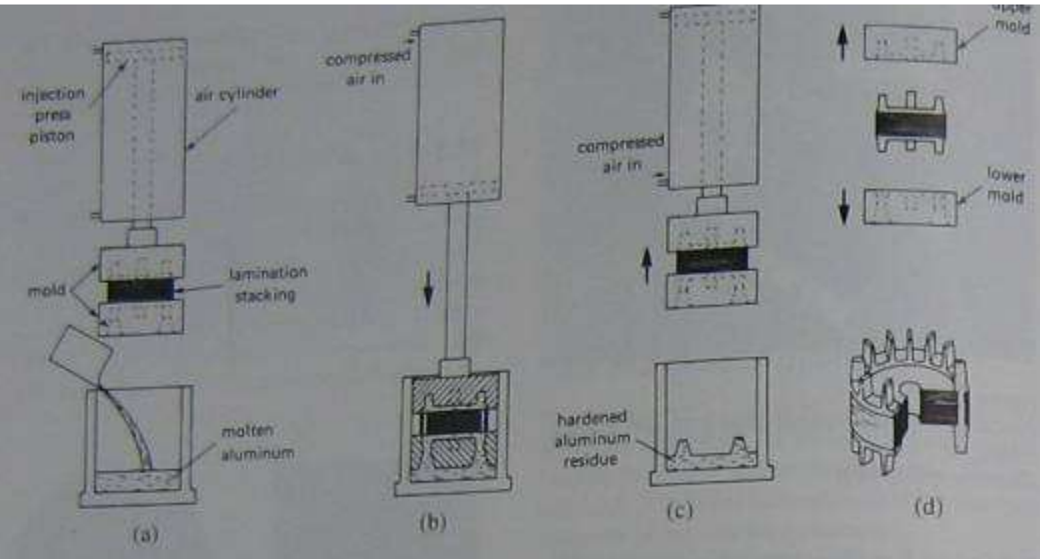


Figure 13.3c
Progressive steps in the injection molding of a squirrel-cage rotor.
a. Molten aluminum is poured into a cylindrical cavity. The laminated rotor stacking is firmly held between two molds.
b. Compressed air rams the mold assembly into the cavity. Molten aluminum is forced upward through the rotor bar holes and into the upper mold.
c. Compressed air withdraws the mold assembly, now completely filled with hot (but hardened) aluminum.
d. The upper and lower molds are pulled away, revealing the die-cast rotor. The cross section view shows that the upper and lower end-rings are joined by the rotor bars. *(Lab-Volt)*

two coils in each winding produce magnetomotive forces that act in the same direction.

The three sets of windings are connected in wye, thus forming a common neutral N. Owing to the perfectly symmetrical arrangement, the line-to-neutral impedances are identical. In other words, as regards terminals A, B, C, the windings constitute a balanced 3-phase system.

If we connect a 3-phase source to terminals A, B, C, alternating currents I_a, I_b, and I_c will flow in the windings. The currents will have the same value but will be displaced in time by an angle of 120°. These currents produce magnetomotive forces which, in turn, create a magnetic flux. It is this flux we are interested in.

In order to follow the sequence of events, we assume that positive currents (indicated by the arrows) always flow in the windings from line to neutral. Conversely, negative currents flow from neutral to line. Furthermore, to enable us to work with numbers, suppose that the peak current per phase is 10 A. Thus, when $I_a = +7$ A, the two coils of phase A will together produce an mmf of 7 A × 10 turns = 70 ampere-turns and a corresponding value of flux. Because the current is positive, the flux is directed vertically upward, according to the right-hand rule.

As time goes by, we can determine the instantaneous value and direction of the current in each winding and thereby establish the successive flux patterns. Thus, referring to Fig. 13.7 at instant 1, current I_a has a value of +10 A, whereas I_b and I_c both have a value of −5 A. The mmf of phase A is 10 A × 10 turns = 100 ampere-turns, while the mmf

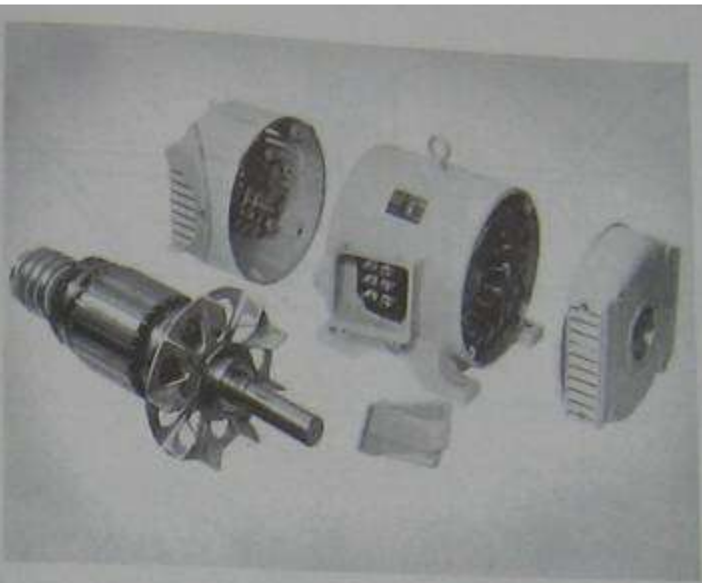


Figure 13.4a
Exploded view of a 5 hp, 1730 r/min wound-rotor induction motor.

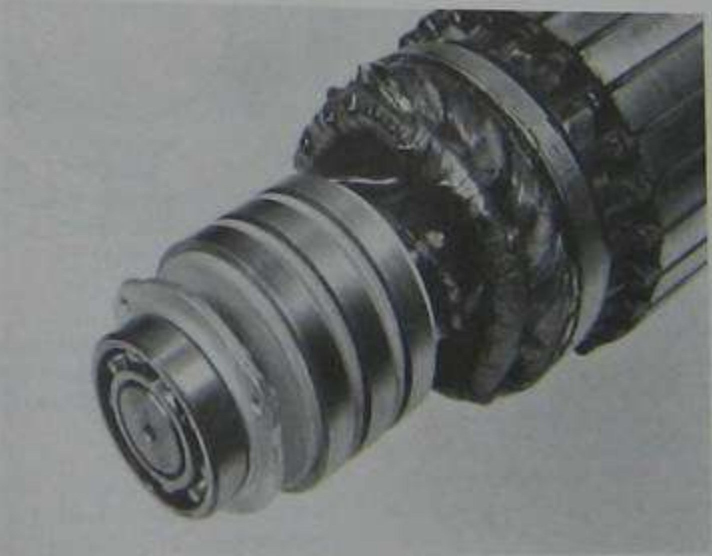


Figure 13.4b
Close-up of the slip-ring end of the rotor.
(Courtesy of Brook Crompton Parkinson Ltd)

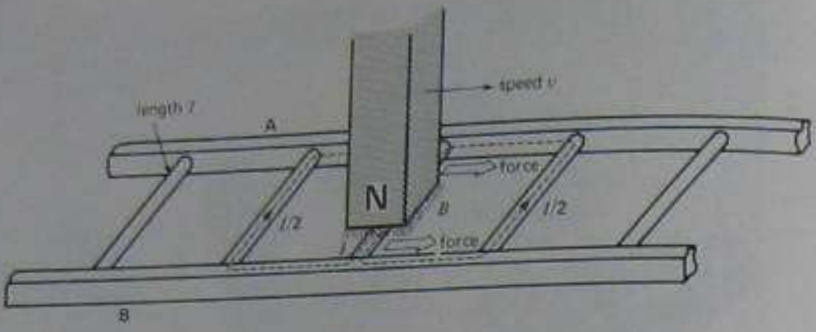


Figure 13.5a
Moving magnet cutting across a conducting ladder.

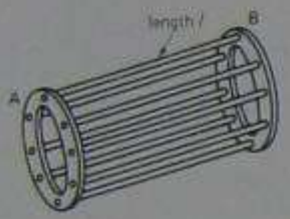


Figure 13.5b
Ladder bent upon itself to form a squirrel-cage.

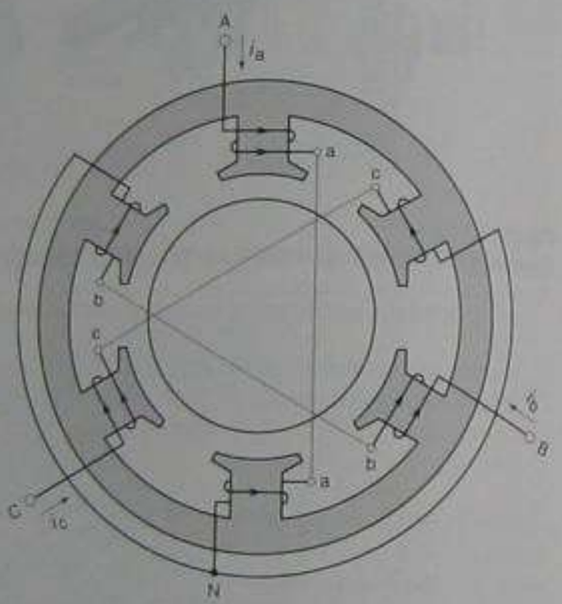


Figure 13.6
Elementary stator having terminals A, B, C connected to a 3-phase source (not shown). Currents flowing from line to neutral are considered to be positive.

of phases B and C are each 50 ampere-turns. The direction of the mmf depends upon the instantaneous current flows and, using the right-hand rule, we find that the direction of the resulting magnetic field is as shown in Fig. 13.8a. Note that as far as the rotor is concerned, the six salient poles together produce a magnetic field having essentially one broad north pole and one broad south pole. This means that the 6-pole stator actually produces a 2-pole field. The combined magnetic field points upward.

At instant 2, one-sixth cycle later, current I_c attains a peak of -10 A, while I_a and I_b both have a value of $+5$ A (Fig. 13.8b). We discover that the new field has the same shape as before, except that it has moved clockwise by an angle of 60°. In other words, the flux makes 1/6 of a turn between instants 1 and 2.

Proceeding in this way for each of the successive instants 3, 4, 5, 6, and 7, separated by intervals of 1/6 cycle, we find that the magnetic field makes one complete turn during one cycle (see Figs. 13.8a to 13.8f).

The rotational speed of the field depends, therefore, upon the duration of one cycle, which in turn depends on the frequency of the source. If the frequency is 60 Hz, the resulting field makes one turn in 1/60 s, that is, 3600 revolutions per minute. On

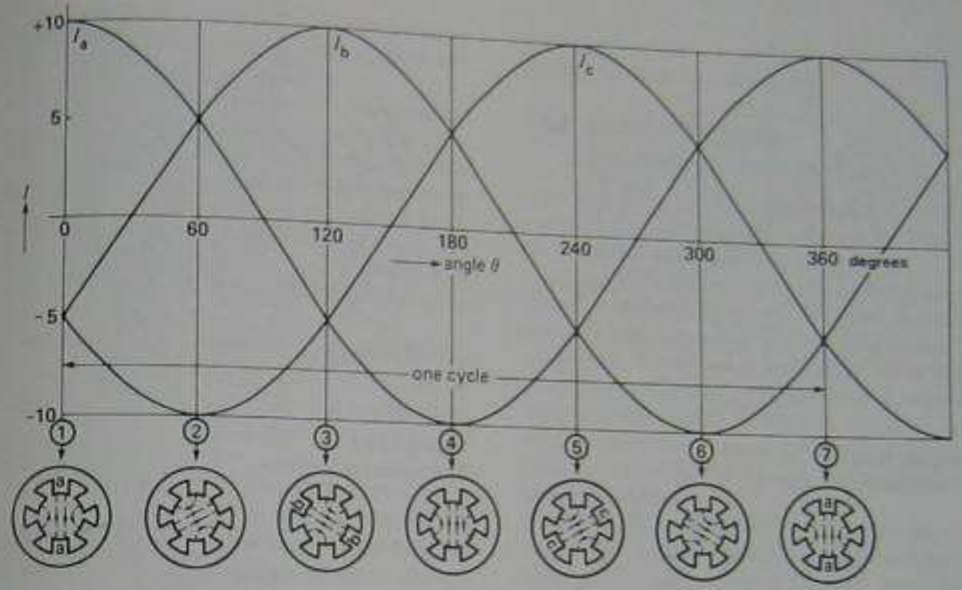


Figure 13.7
Instantaneous values of currents and position of the flux in Fig. 13.6.

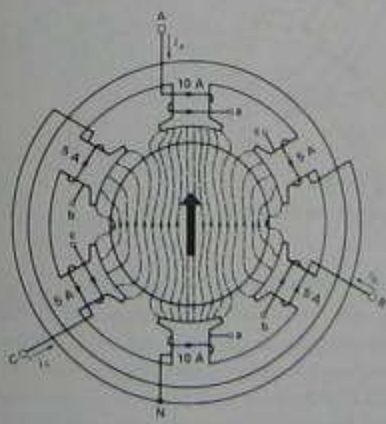


Figure 13.8a
Flux pattern at instant 1.

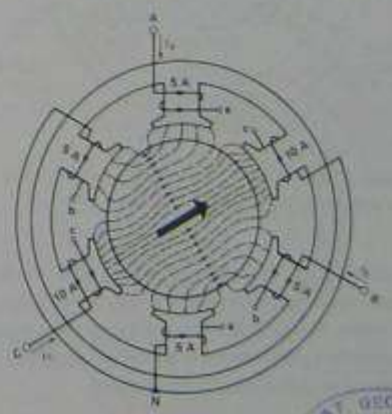


Figure 13.8b
Flux pattern at instant 2.

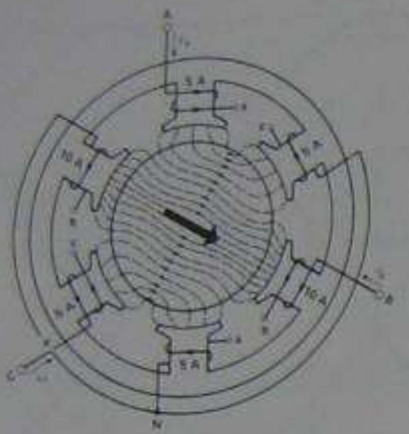


Figure 13.8c
Flux pattern at instant 3.

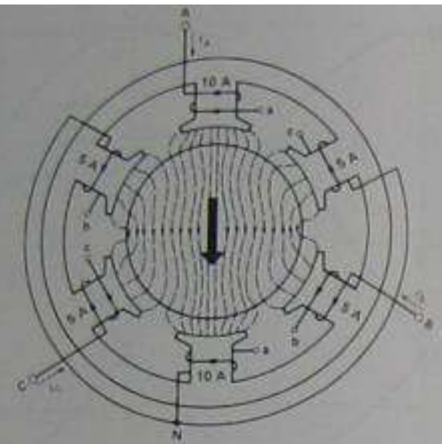


Figure 13.8d
Flux pattern at instant 4.

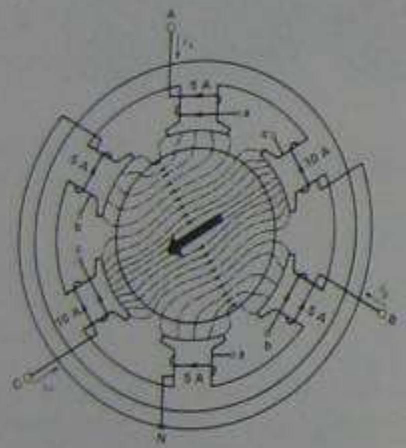


Figure 13.8e
Flux pattern at instant 5.

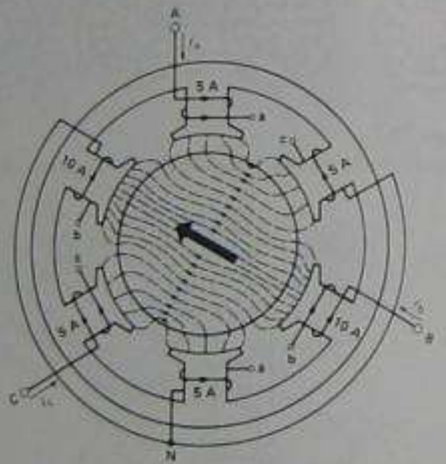


Figure 13.8f
Flux pattern at instant 6.

the other hand, if the frequency were 5 Hz, the field would make one turn in 1/5 s, giving a speed of only 300 r/min. Because the speed of the rotating field is necessarily synchronized with the frequency of the source, it is called *synchronous speed.*

13.4 Direction of rotation

The positive crests of the currents in Fig. 13.7 follow each other in the order A-B-C. This phase sequence produces a field that rotates clockwise. If we interchange any two of the lines connected to the stator, the new phase sequence will be A-C-B. By following the same line of reasoning developed in Section 13.3, we find that the field now revolves at synchronous speed in the opposite, or counterclockwise direction. Interchanging any two lines of a 3-phase motor will, therefore, reverse its direction of rotation.

Although early machines were built with salient poles, the stators of modern motors have internal diameters that are smooth. Thus, the salient-pole stator of Fig. 13.6 is now replaced by a smooth stator such as shown in Figs. 13.2 and 13.24a.

In Fig. 13.6, the two coils of phase A (Aa and An) are replaced by the two coils shown in Fig. 13.9a. They are lodged in two slots on the inner surface of the stator. Note that each coil covers 180° of the circumference whereas the coils in Fig. 13.6 cover only 60°. The 180° coil pitch is more efficient because it produces more flux per turn. A current I_a flowing from terminal A to the neutral N yields the flux distribution shown in the figure.

The coils of phases B and C are identical to those of phase A and, as can be seen in Fig. 13.9b, they are displaced at 120° to each other. The resulting magnetic field due to all three phases again consists of two poles.

In practice, instead of using a single coil per pole as shown in Fig. 13.9a, the coil is subdivided into two, three or more coils lodged in adjacent slots. The staggered coils are connected in series and constitute what is known as a *phase group*. Spreading the coil in this way over two or more slots tends to create a sinusoidal flux distribution per pole, which improves the performance of the motor and makes it less noisy. A phase group (or simply *group*) composed of 5 staggered coils connected in series to be placed in 5 successive slots is shown in Fig. 13.20.

13.5 Number of poles—synchronous speed

Soon after the invention of the induction motor, it was found that the speed of the revolving flux could be reduced by increasing the number of poles.

To construct a 4-pole stator, the coils are distributed as shown in Fig. 13.10a. The four identical groups of phase A now span only 90° of the stator circumference. The groups are connected in series and in such a way that adjacent groups produce magnetomotive forces acting in opposite directions. In other words, when a current I_a flows in the stator winding of phase A (Fig. 13.10a), it creates four alternate N-S poles.

The windings of the other two phases are identical but are displaced from each other (and from phase A) by a mechanical angle of 60°. When the wye-connected windings are connected to a 3-phase source, a revolving field having four poles is created (Fig. 13.10b). This field rotates at only half the speed of the 2-pole field shown in Fig. 13.9b. We will shortly explain why this is so.

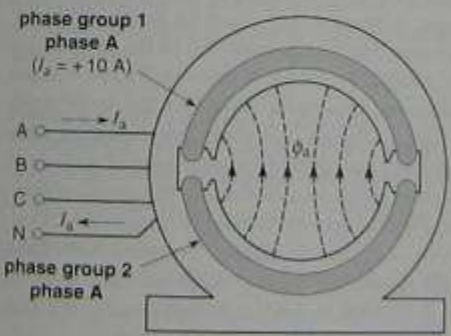


Figure 13.9a
Phase group 1 is composed of a single coil lodged in two slots. Phase group 2 is identical to Phase group 1. The two coils are connected in series. In practice, a phase group usually consists of two or more staggered coils.

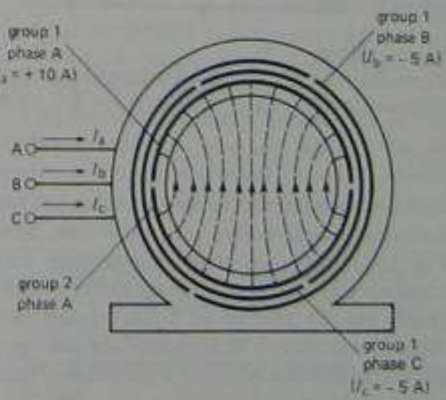


Figure 13.9b
Two-pole, full-pitch, lap-wound stator and resulting magnetic field when the current in phase A = +10 A and $I_b = I_c = -5$ A.

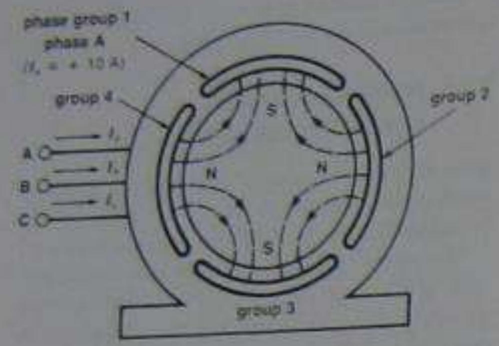


Figure 13.10a
The four phase groups of phase A produce a 4-pole magnetic field.

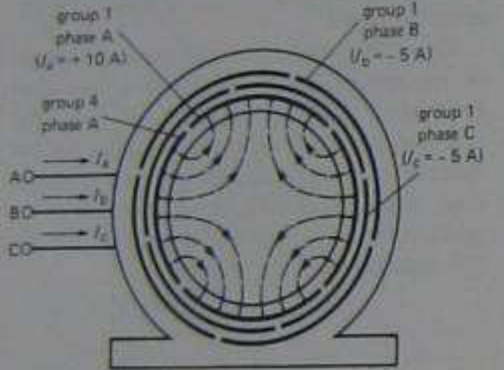


Figure 13.10b
Four-pole, full-pitch, lap-wound stator and resulting magnetic field when $I_a = +10$ A and $I_b = I_c = -5$ A.

We can increase the number of poles as much as we please provided there are enough slots. Thus, Fig. 13.11 shows a 3-phase, 8-pole stator. Each phase consists of 8 groups, and the groups of all the phases together produce an 8-pole rotating field. When connected to a 60 Hz source, the poles turn, like the spokes of a wheel, at a synchronous speed of 900 r/min.

How can we tell what the synchronous speed will be? Without going into all the details of current flow in the three phases, let us restrict our attention to phase A. In Fig. 13.11 each phase group covers a mechanical angle of 360/8 = 45°. Suppose the current in phase A is at its maximum positive value. The magnetic flux is then centered on phase A, and the N-S poles are located as shown in Fig. 13.12a. One-half cycle later, the current in phase A will reach its maximum negative value. The flux pattern will be the same as before, except that all the N poles will have become S poles and vice versa (Fig. 13.12b). In comparing the two figures, it is clear that the entire magnetic field has shifted by an angle of 45°—and this gives us the clue to finding the speed of rotation. The flux moves 45° and so it takes 8 half-cycles (= 4 cycles) to make a complete turn. On a 60 Hz system the time to make one turn is therefore $4 \times 1/60 = 1/15$ s. Consequently, the flux turns at the rate of 15 r/s or 900 r/min.

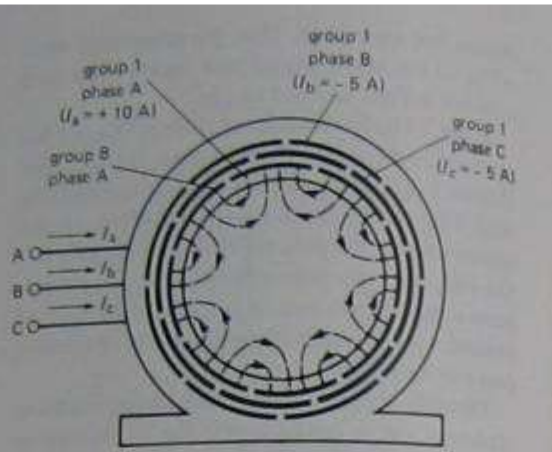


Figure 13.11
Eight-pole, full-pitch, lap-wound stator and resulting magnetic field when $I_a = +10$ A and $I_b = I_c = -5$ A.

The speed of a rotating field depends therefore upon the frequency of the source and the number of poles on the stator. Using the same reasoning as above, we can prove that the synchronous speed is always given by the expression

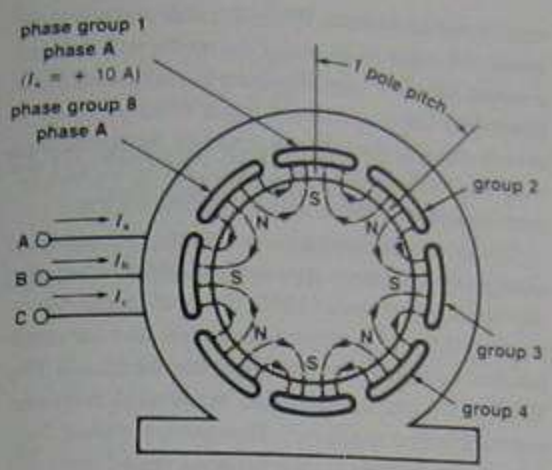


Figure 13.12a
Flux pattern when the current in phase A is at its maximum positive value.

$$n_s = \frac{120f}{p} \qquad (13.1)$$

where

n_s = synchronous speed [r/min]
f = frequency of the source [Hz]
p = number of poles

This equation shows that the synchronous speed increases with frequency and decreases with the number of poles.

Example 13-1

Calculate the synchronous speed of a 3-phase induction motor having 20 poles when it is connected to a 50 Hz source.

Solution

$$n_s = 120f/p = 120 \times 50/20$$
$$= 300 \text{ r/min}$$

Learning Outcome 1.3

Determine the volts per turn, pitch and breadth factors given the desired value of stator flux density

THREE-PHASE WINDINGS AND FIELDS

In an a.c. machine the armature (or main) winding may be either on the stator (i.e. the stationary part of the machine) or on the rotor, the same form of winding being used in each case. The simplest form of 3-phase winding has concentrated coils each spanning one pole pitch, and with the starts of each spaced 120° (electrical) apart on the stator or rotor. These coils may be connected in star or delta as required.

In most machines the coils are not concentrated but are distributed in slots over the surface of the stator or rotor, and it is this type of winding which will now be considered. The same type of winding is common to both synchronous and asynchronous (induction) machines.

11.1 Flux Density Distributions

In all a.c. machines an attempt is made to secure a sinusoidal flux density distribution in the air-gap. This may be achieved approximately by the distribution of the winding in slots round the air-gap or by using salient poles with shaped pole shoes.

In Fig. 11.1(*a*) a section of a multipolar machine is shown. If the flux density in the air-gap is to be sinusoidally distributed, the flux density must be zero on the inter-polar axes such as OA, OC and OE, and maximum on the polar axes OB and OD. Since successive poles are of alternate north and south polarities, the maximum flux densities along OB and OD are oppositely directed. Thus a complete cycle of variation of the flux density takes place in a

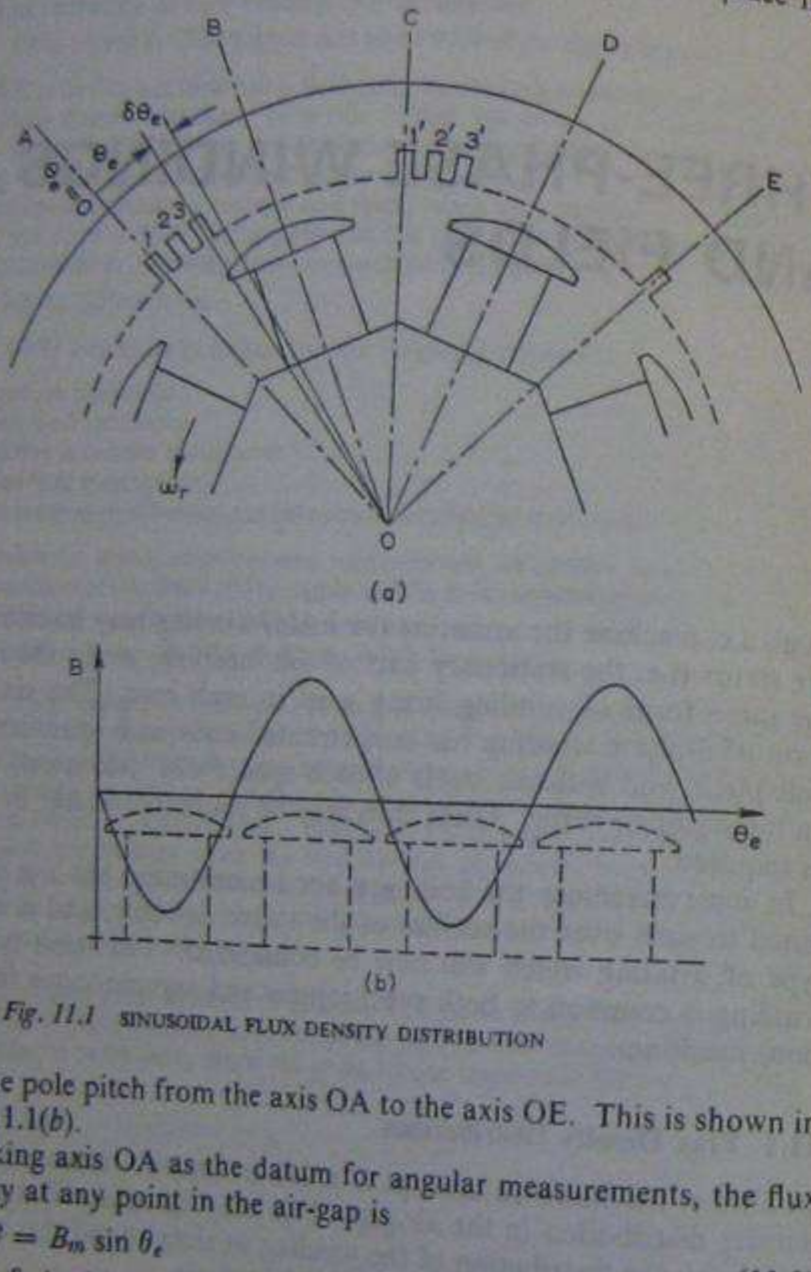


Fig. 11.1 SINUSOIDAL FLUX DENSITY DISTRIBUTION

double pole pitch from the axis OA to the axis OE. This is shown in Fig. 11.1(*b*).

Taking axis OA as the datum for angular measurements, the flux density at any point in the air-gap is

$$B = B_m \sin \theta_e \tag{11.1}$$

where θ_e is the angle from the origin measured in electrical radians or electrical degrees. Since one cycle of variation of the flux density occurs in a double pole pitch,

1 double pole pitch $\equiv 2\pi$ electrical radians or 360 electrical degrees

If the machine has $2p$ poles or p double pole pitches,

$$\theta_e = p\theta_m \tag{11.2}$$

where θ_m is the angular measure in mechanical radians or degrees.

11.2 Three-phase Single-layer Concentric Windings

The two sides of an armature coil must be placed in slots which are approximately a pole pitch (180 electrical degrees) apart so that the e.m.f.s in the coil sides are cumulative. In addition, in 3-phase machines the starts of each phase winding must be 120 electrical degrees apart.

In single-layer windings one coil side occupies the whole of a slot. As a result, difficulty is experienced in arranging the end connectors, or overhangs. In concentric and split-concentric windings differently shaped coils having different spans are necessary. To preserve e.m.f. balance in each of the phases, each phase must contain the same number of each shape of coil.

Fig. 11.2(*a*) represents a developed stator with 24 stator slots, and it is desired to place a 4-pole 3-phase concentric winding in them:

$$\text{Number of slots per pole} = \frac{24}{4} = 6$$

$$\text{Number of slots per pole and phase} = \frac{24}{4 \times 3} = 2$$

Fig. 11.2(*a*) shows the coil arrangement for the red phase as a thin full line. The start and finish (marked S and F respectively) of the phase winding are brought out, all the coils in the one phase being connected in series. For a phase sequence RYB, the yellow phase (shown dotted) must start 120 electrical degrees after the red phase. One pole pitch contains six slots and is equivalent to 180 electrical degrees. Hence a slot pitch is equivalent, in this case, to 30 electrical degrees.

The red phase starts in slot 1 and therefore the yellow phase must start in slot 5. In the same way the blue phase is 240 electrical degrees out of space phase with the red phase. The blue phase must therefore start in slot 9.

In Fig. 11.2 the finishes of the three phases have been commoned, making a star-connected winding. It would have been equally correct to common the three starts. The winding might also have been mesh-connected, in which case the finish of the red phase would have been connected to the start of the yellow phase, the finish of the yellow to the start of the blue, the finish of the blue to the start of the red, three connectors to the three junctions being brought out to terminals.

It will be observed that each phase has coils of each of the four different sizes used, thus maintaining balance between the phases.

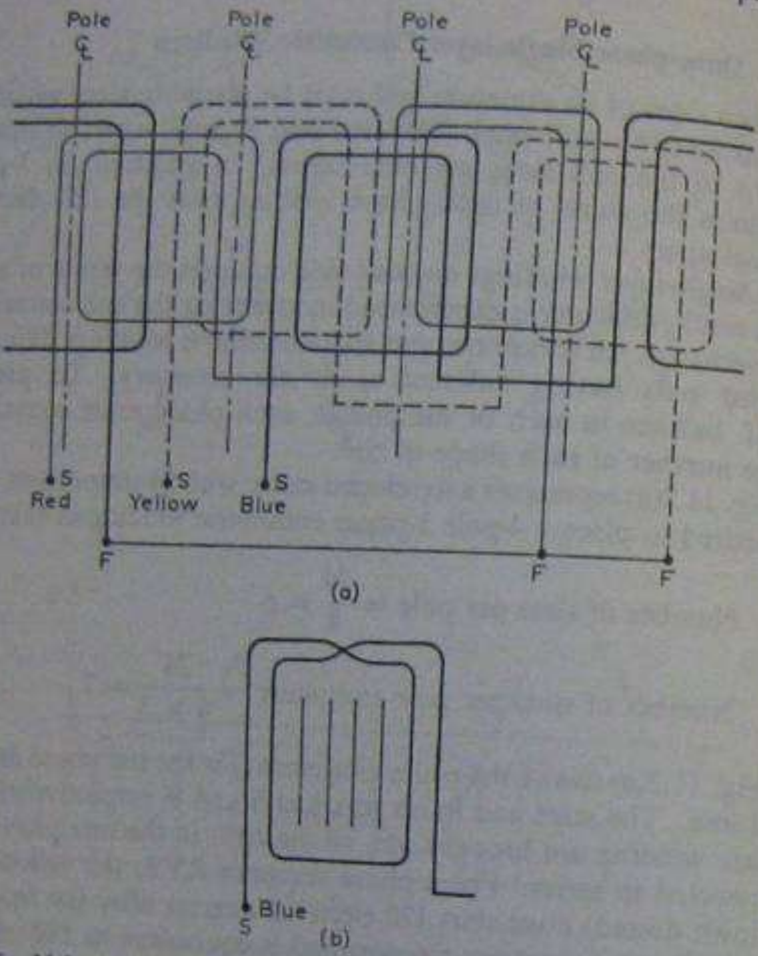


Fig. 11.2 FOUR-POLE 3-PHASE SINGLE-LAYER CONCENTRIC WINDING

It will also be seen that a coil group of any one phase consists of two coils per double pole pitch, one coil being greater than a pole pitch by one slot pitch and the other being less than a pole pitch by the same amount. If the end connexions of these two coils were crossed over as shown in Fig. 11.2(*b*) two full-pitch coils (i.e. having a span of exactly one-pole pitch) would be formed. Therefore each such coil group is the equivalent, electrically, of two full-pitch coils joined in series. All single-layer windings are effectively composed of full-pitch coils.

11.3 Three-phase Single-layer Mush Winding

Fig. 11.3 shows a 4-pole 3-phase single-layer mush winding. The distinctive feature of the mush winding is the utilization of constant-span coils. The overhangs are arranged in a similar manner to those of a conventional double-layer winding.

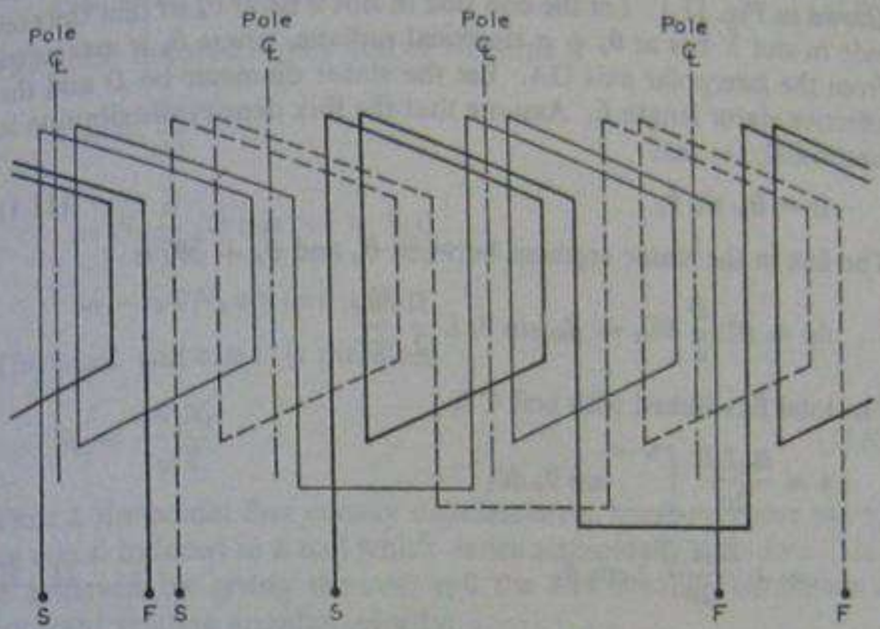


Fig. 11.3 FOUR-POLE 3-PHASE SINGLE-LAYER MUSH WINDING

11.4 Three-phase Double-layer Windings

The double-layer windings used in 3-phase machines are essentially similar to those used in d.c. machines except that no connexions to a commutator are required.

Since each phase must be balanced, all must contain equal numbers of coils and the starts of each phase must be displaced by 120 electrical degrees. If a number of groups of coils are to be connected in parallel, then similar parts in the winding at equal potentials must be available, a condition obtainable only in machines having a number of poles divisible by three when a wave winding is used.

On the other hand, tooth ripple, which arises where there are an integral number of slots per pole, resulting in the same relative positions of equivalent slots under each pole, may be avoided in double-layer windings by the use of winding pitches different from the pole pitch, thus giving a fractional number of slots per pole. A further advantage of the double-layer winding is the possibility of using constant-span coils. Only single-layer windings are considered in the rest of this chapter.

11.5 E.M.F. Induced in a Full-pitch Coil

Consider a full-pitch coil C with coil sides lying in slots 3 and 3′ as shown in Fig. 11.1. Let the coil side in slot 3 lie at θ_e so that the coil side in slot 3′ lies at $\theta_e + \pi$ electrical radians, where θ_e is measured from the interpolar axis OA. Let the stator diameter be D and the effective stator length L. Assume that the flux density distribution is sinusoidal, i.e. that

$$B = B_m \sin \theta_e \tag{11.1}$$

The flux in the stator segment between θ_e and $\theta_e + \delta\theta_e$ is

$$\delta\phi = BL\frac{D}{2}\,\delta\theta_m = B_m \sin \theta_e L \frac{D}{2}\frac{\delta\theta_e}{p}$$

The total flux linked with coil C is

$$\phi = \frac{B_m LD}{2p}\int_{\theta_e}^{\theta_e+\pi} \sin \theta_e \, d\theta_e$$

$$= +\frac{B_m LD}{2p} 2 \cos \theta_e \tag{11.3}$$

If a coil lies with its sides on the interpolar axes, as, for example, the coil lying in slots 1 and 1′ of Fig. 11.1, then the coil links the total flux per pole, Φ:

$$\Phi = \frac{B_m LD}{2p}\int_0^{\pi} \sin \theta_e \, d\theta_e$$

$$= +\frac{B_m LD}{2p} 2 \tag{11.4}$$

The flux linked with coil C is therefore, by substitution in eqn. (11.3),

$$\phi = \Phi \cos \theta_e \tag{11.5}$$

Suppose the pole system rotates in the direction shown at a uniform angular velocity

$$\omega_r = 2\pi n_0 \quad \text{radians/second} \tag{11.6}$$

where n_0 is the rotor speed in revolutions per second. The position of any coil such as C at any instant, in electrical radians, is

$$\theta_e = \omega t + \theta_0$$

where θ_0 is the position of the coil at $t = 0$, and

$$\omega = p\omega_r = 2\pi n_0 p \quad \text{electrical radians/second} \tag{11.7}$$

Substituting for θ_e in eqn. (11.5), the flux linking any coil such as C at any time t is

$$\phi = \Phi \cos(\omega t + \theta_0) \tag{11.8}$$

The e.m.f. induced in any coil of N_c turns is

$$\begin{aligned} e &= N_c \frac{d\phi}{dt} \\ &= N_c \frac{d}{dt}\{\Phi \cos(\omega t + \theta_0)\} \\ &= -\omega\Phi N_c \sin(\omega t + \theta_0) \end{aligned}$$

The r.m.s. coil e.m.f. is therefore

$$E_c = \frac{\omega\Phi N_c}{\sqrt{2}} \tag{11.9}$$

Thus a sinusoidal flux density distribution in space may give rise to an e.m.f. induced in a coil which varies sinusoidally with time. This is achieved by giving the coil and the flux density distribution a constant relative angular velocity.

The frequency of the induced e.m.f. is

$$f = \frac{\omega}{2\pi} = \frac{2\pi n_0 p}{2\pi} = n_0 p \tag{11.10}$$

n_0 is called the *synchronous speed*. In this equation it is measured in revolutions per second.

11.6 Distribution (or Breadth) Factor and E.M.F. Equation

Suppose that under each pole pair each phase of the winding has g coils connected in series, each coil side being in a separate slot. The e.m.f. per phase and pole pair is the complexor sum of the coil voltages. These will not be in time phase with one another since successive coils are displaced round the armature, and hence will not be linked by the same value of flux at the same instant. $\boldsymbol{E}_1$, $\boldsymbol{E}_2$, $\boldsymbol{E}_3$ . . . $\boldsymbol{E}_g$ (as shown in Fig. 11.4(a)) represent the r.m.s. values of the e.m.f.s in successive coils. The phase displacement between successive e.m.f.s is ψ, which depends on the electrical angular displacement between successive slots on the armature.

Suppose the machine has a total of S slots and $2p$ poles. Then

$$\text{Number of slots per pole} = \frac{S}{2p}$$

The slot pitch (electrical angle between slot centre lines) is

$$\psi = \frac{180°_e}{S/2p} \quad (\text{since 1 pole pitch} = 180°_e) \tag{11.11}$$

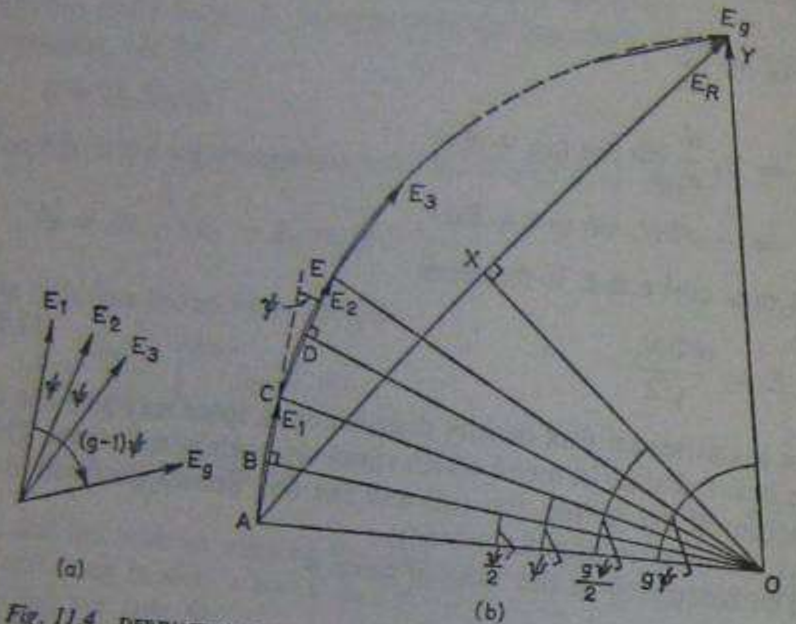


Fig. 11.4 DERIVATION OF DISTRIBUTION FACTOR
(a) Complexor diagram of slot e.m.f.s
(b) Resultant of slot e.m.f.s

The e.m.f. complexors $E_1, E_2, E_3 \ldots E_g$ are placed end to end in order in Fig. 11.4(*b*). The resultant complexor E_R, represents the complexor sum of the e.m.f.s of the g coils connected in series.

Since the complexors $E_1, E_2, E_3 \ldots E_g$ are all of the same length and are displaced from one another by the same angle, they must be successive chords of the circle whose centre is O in Fig. 11.4(*b*). The complexor sum AY may be found as follows.

Join OA, OC, OE, etc., draw the perpendicular bisectors of each chord (i.e. OB, OD, etc.) and also the perpendicular bisector OX of the chord AY.

In the triangle AOX,

$$AX = AO \sin AOX = AO \sin g\frac{\psi}{2}$$

Therefore

$$AY = 2AO \sin g\frac{\psi}{2}$$

In the triangle AOB,

$$AB = AO \sin AOB = AO \sin\frac{\psi}{2}$$

$$AC = 2AB = 2AO \sin\frac{\psi}{2}$$

Therefore

$$\frac{AY}{AC} = \frac{E_R}{E_1} = \frac{\sin g\frac{\psi}{2}}{\sin\frac{\psi}{2}}$$

Thus the *distribution factor* is

$$K_d = \frac{\text{Complexor sum of coil e.m.f.s}}{\text{Arithmetic sum of coil e.m.f.s}}$$

$$= \frac{E_R}{gE_1} = \frac{\sin g\frac{\psi}{2}}{g \sin\frac{\psi}{2}} \qquad (11.12)$$

The product $g\psi$ represents the electrical angle over which the conductors of one phase are spread under any one pole and is referred to as the *phase spread*. In a 3-phase single-layer winding each phase has two phase spreads under each pole pair. Therefore, for a single-layer 3-phase winding,

$$g\psi = \frac{360}{2 \times 3} = 60°_e \quad \text{or} \quad \pi/3 \text{ electrical radians}$$

Clearly the highest value which the distribution factor K_d can have is unity, corresponding to a situation where there is one coil per pole pair and phase. A lower limit for the value of K_d also exists. Thus, if the number of separate slots g in the phase spread $g\psi$ is considered to increase without limit, then

$$\psi \to 0 \quad \text{and} \quad \sin\frac{\psi}{2} \to \frac{\psi}{2}$$

A 3-phase winding with a phase spread of $60°_e$ is said to be *narrow spread*.

For a narrow-spread 3-phase winding ($g\psi = \pi/3$),

$$\lim_{\psi \to 0} K_d = \frac{\sin \frac{g\psi}{2}}{g\frac{\psi}{2}} = \frac{\sin \pi/6}{\pi/6} = \frac{3}{\pi} \tag{11.13}$$

A winding having this limiting condition is called a *uniform winding*, and in such winding the phase spreads may be thought of as current sheets with the effect of the slotting eliminated.

The lower limit of K_d for a 3-phase narrow-spread winding ($3/\pi = 0{\cdot}955$), corresponding to a very large number of slots per pole and phase, shows that the distribution of the winding will have little effect on the magnitude of the fundamental e.m.f. per phase.

Ideally the flux density distribution linking the winding should be sinusoidal. In practice this ideal is not usually achieved; the air-gap flux density distribution is then of the form

$$B = B_{m1} \sin \theta_e + B_{m3} \sin (3\theta_e + \epsilon_3) + \ldots B_{mn} \sin (n\theta_e + \epsilon_n) \tag{11.14}$$

In this expression the first term on the right-hand side is called the *fundamental space distribution*. The other terms are referred to as *space harmonics*. The *n*th space harmonic goes through *n* cycles of variation for one cycle of variation of the fundamental. Only odd space harmonics are present since the flux density distribution repeats itself under each pole and is therefore symmetrical.

Just as the fundamental flux density gives rise to a fundamental e.m.f. induced in a coil, so the *n*th space harmonic in the flux density distribution will give rise to an *n*th time harmonic in the coil e.m.f. The distribution factor for the *n*th harmonic is

$$K_{dn} = \frac{\sin \frac{gn\psi}{2}}{g \sin \frac{n\psi}{2}} \tag{11.15}$$

Although the distribution of the winding has little effect on the magnitude of the fundamental, it may cause considerable reduction in the magnitude of harmonic e.m.f.s compared with those occurring in a winding for which $g = 1$, i.e. one coil per pole pair and phase.

11.7 Coil-span Factor

The e.m.f. equation of Section 11.5 has been deduced on the assumption of full-pitch coils, i.e. coils whose sides are separated by one pole pitch. As has been pointed out, the coils in double-layer windings are often made either slightly more or slightly less than a pole pitch. Fig. 11.5 illustrates coils with various pitches.

If the coil has a pitch of exactly one pole pitch, it will at some instant link the entire flux of a rotor pole. If the coil pitch is less than one pole pitch, it will never link the entire flux of a rotor pole and the maximum coil e.m.f. will be reduced. If the coil pitch is greater than one pole pitch, the coil must always be linking flux

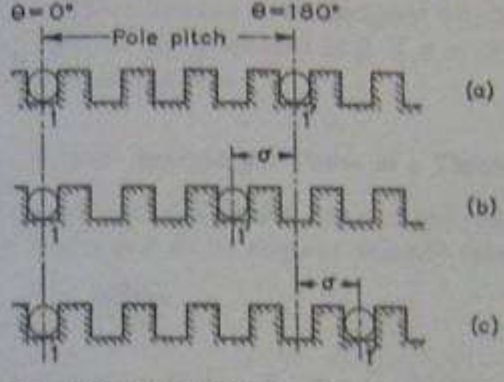


Fig. 11.5 COIL SPANS
(*a*) Full pitch
(*b*) Short pitch
(*c*) Over-full pitch

from at least two adjacent rotor poles so that the net flux linked will be less than the flux of one pole and the maximum coil e.m.f. will again be reduced.

The factor by which the e.m.f. per coil is reduced is called the *coil span factor*, K_s:

$$K_s = \frac{\text{E.M.F. in the short or long coil}}{\text{E.M.F. in a full-pitched coil}} \qquad (11.16)$$

The magnitude of the coil span factor may most readily be obtained by considering the e.m.f. induced in each coil side, namely

$$e = Blv \text{ volts}$$

where B = air-gap flux density, l = active conductor length and v = conductor velocity at right angles to the direction of B.

This e.m.f. will have the same waveform as the flux density in the air-gap, since l and v are constant, and hence if the flux density is sinusoidally distributed the e.m.f. in each conductor will be sinusoidal so that the resultant coil e.m.f. will also be sinusoidal. If the pitch is short or long by an electrical angle σ, then, assuming a sinusoidal flux density distribution, the e.m.f.s in each side of the coil will differ in phase by σ but will have the same r.m.s. value. The resultant coil e.m.f. will be the complexor sum of the e.m.f.s in each coil side, as shown in Fig. 11.6.

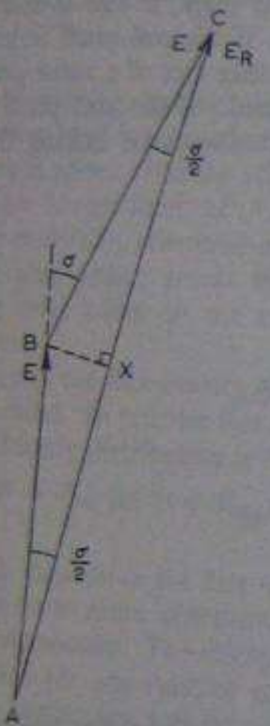


Fig. 11.6 DERIVATION OF COIL SPAN FACTOR

$$\text{Resultant e.m.f.} = \text{AC} = 2\text{AB}\cos\frac{\sigma}{2}$$

$$\text{E.M.F. for a full-pitch coil} = 2\text{AB}$$

Therefore

$$K_s = \frac{2\text{AB}\cos\frac{\sigma}{2}}{2\text{AB}} = \cos\frac{\sigma}{2} \tag{11.17}$$

If the flux density distribution contains space harmonics, the coil span factor for the nth harmonic e.m.f. is

$$K_{sn} = \cos\frac{n\sigma}{2} \tag{11.18}$$

All single-layer windings are effectively made up of full-pitch coils, but double-layer windings usually have short-pitched or short-chorded coils. The nth harmonic coil e.m.f. is reduced to zero if the *chording angle*, σ, is such that

$$\cos\frac{n\sigma}{2} = 0$$

or

$$\frac{n\sigma}{2} = 90°_e \tag{11.19}$$

This enables windings to be designed which will not permit specified harmonics to be generated (e.g. if $\sigma = 60°_e$ there can be no third-harmonic generation).

11.8 E.M.F. Induced per Phase of a Three-phase Winding

Following eqn. (11.9) the r.m.s. e.m.f. induced in a full-pitch coil of N_c turns due to its angular velocity relative to the pole system is

$$E_c = \frac{\omega\Phi N_c}{\sqrt{2}} \tag{11.9}$$

For a coil-span factor, K_s, due to chording,

$$E_c = K_s\frac{\omega\Phi N_c}{\sqrt{2}}$$

Further, if there are g coils in a phase group under a pole pair the resultant complexor sum is

$$E_g = K_d g E_c = K_d K_s g\frac{\omega\Phi N_c}{\sqrt{2}}$$

Assuming that the e.m.f.s of coil groups of the same phase under successive pole pairs are in phase and connected in series, the e.m.f. per phase is

$$E_p = pEg = pK_d K_s g\frac{\omega\Phi N_c}{\sqrt{2}}$$

or

$$E_p = K_d K_s\frac{\omega\Phi N_p}{\sqrt{2}} \tag{11.20}$$

where the number of turns per phase, N_p, is pgN_c.

This equation is sometimes written in the form

$$E_p = 4{\cdot}44\,K_d K_s f\Phi N_p \tag{11.21}$$

since $\omega = 2\pi f$ and $2\pi/\sqrt{2} = 4{\cdot}44$.

Learning Outcome 1.4

Determine the rotor impedance, current and power factor at a given value of slip

Example 13-6

A 3-phase, 8-pole squirrel-cage induction motor, connected to a 60 Hz line, possesses a synchronous speed of 900 r/min. The motor absorbs 40 kW, and the copper and iron losses in the stator amount to 5 kW and 1 kW, respectively. Calculate the torque developed by the motor.

Solution

The power transmitted across the air gap to the rotor is

$$P_r = P_e - P_{js} - P_f$$

$$= 40 - 5 - 1 = 34 \text{ kW}$$

$$T_m = 9.55\, P_r / n_s \qquad (13.9)$$

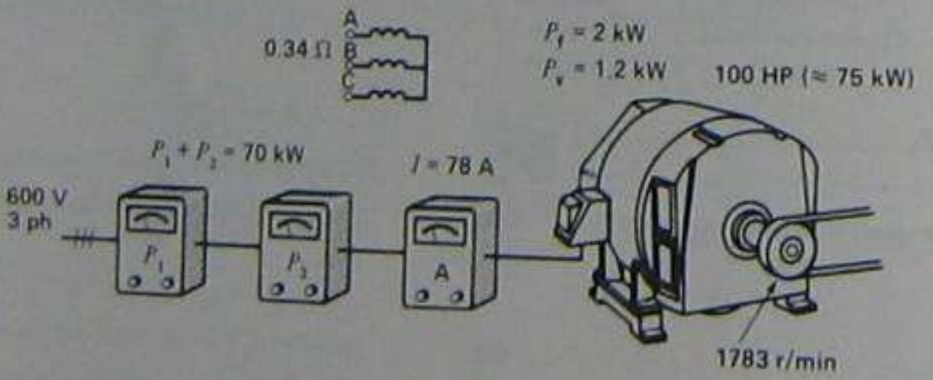
$$= 9.55 \times 34\,000/900$$

$$= 361 \text{ N·m}$$

Note that the solution to this problem (the torque) is independent of the speed of rotation. The motor could be at a standstill or running at full speed, but as long as the power P_r transmitted to the rotor is equal to 34 kW, the motor develops a torque of 361 N·m.

Example 13-7

A 3-phase induction motor having a nominal rating of 100 hp (~75 kW) and a synchronous speed of 1800 r/min is connected to a 600 V source (Fig. 13.16a). The two-wattmeter method shows a total power contwo-wattmeter method shows a total pow



6a
le 13-7.

sumption of 70 kW, and an ammeter indicates a line current of 78 A. Precise measurements give a rotor speed of 1763 r/min. In addition, the following characteristics are known about the motor:

stator iron losses $P_f = 2$ kW
windage and friction losses $P_v = 1.2$ kW
resistance between two stator terminals $= 0.34\ \Omega$

Calculate

a. Power supplied to the rotor
b. Rotor I^2R losses
c. Mechanical power supplied to the load, in horsepower
d. Efficiency
e. Torque developed at 1763 r/min

Solution

a. Power supplied to the stator is

$$P_e = 70 \text{ kW}$$

Stator resistance per phase (assume a wye connection) is

$$R = 0.34/2 = 0.17\ \Omega$$

Stator I^2R losses are

$$P_{js} = 3\,I^2R = 3 \times (78)^2 \times 0.17 = 3.1 \text{ kW}$$

Iron losses $P_f = 2$ kW
Power supplied to the rotor:

$$P_r = P_e - P_{js} - P_f = (70 - 3.1 - 2) = 64.9 \text{ kW}$$

b. The slip is

$$s = (n_s - n)/n_s = (1800 - 1763)/1800 = 0.0205$$

Rotor I^2R losses:

$$P_{jr} = sP_r = 0.0205 \times 64.9 = 1.33 \text{ kW}$$

c. Mechanical power developed is

$$P_m = P_r - P_{jr} = 64.9 - 1.33 = 63.5 \text{ kW}$$

Mechanical power P_L to the load:

$$P_L = 63.5 - P_v = 63.5 - 1.2 = 62.3 \text{ kW} = 62.3 \times 1.34 \text{ (hp)} = 83.5 \text{ hp}$$

d. Efficiency of the motor is

$$\eta = P_L/P_e = 62.3/70 = 0.89 \text{ or } 89\%$$

e. Torque at 1763 r/min:

$$T = 9.55\,P_r/n_s = 9.55 \times 64\,900/1800 = 344 \text{ N·m}$$

The above calculations are summarized in Fig. 13.16b.

13.14 Torque versus speed curve

The torque developed by a motor depends upon its speed, but the relationship between the two cannot be expressed by a simple equation. Consequently, we prefer to show the relationship in the form of a

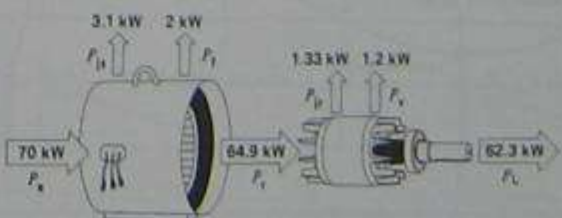


Figure 13.16b
Power flow in Example 13-7.

curve. Fig. 13.17 shows the torque-speed curve of a conventional 3-phase induction motor whose nominal full-load torque is T. The starting torque is 1.5 T and the maximum torque (called *breakdown torque*) is 2.5 T. Pull-up torque is the minimum torque developed by the motor while it is accelerating from rest to the breakdown torque.

At full-load the motor runs at a speed n. If the mechanical load increases slightly, the speed will drop until the motor torque is again equal to the load torque. As soon as the two torques are in balance, the motor will turn at a constant but slightly lower speed. However, if the load torque exceeds 2.5 T (the breakdown torque), the motor will quickly stop.

Small motors (15 hp and less) develop their breakdown torque at a speed n_d of about 80% of synchronous speed. Big motors (1500 hp and more) attain their breakdown torque at about 98% of synchronous speed.

13.15 Effect of rotor resistance

The rotor resistance of a squirrel-cage rotor is essentially constant from no-load to full-load, except that it increases with temperature. Thus, the resistance increases with increasing load because the temperature rises.

In designing a squirrel-cage motor, the rotor resistance can be set over a wide range by using copper, aluminum, or other metals in the rotor bars and end-rings. The torque-speed curve is greatly affected by such a change in resistance. The only characteristic that remains unchanged is the breakdown torque. The following example illustrates the changes that occur.

Figure 13.18a shows the torque-speed curve of a 10 kW (13.4 hp), 50 Hz, 380 V motor having a synchronous speed of 1000 r/min and a full-load torque of 100 N·m (~73.7 ft·lbf). The full-load current is 20 A and the locked-rotor current is 100 A. The rotor has an arbitrary resistance R.

Let us increase the rotor resistance by a factor of 2.5. This can be achieved by using a material of higher resistivity, such as bronze, for the rotor bars and end-rings. The new torque-speed curve is shown in Figure 13.18b. It can be seen that the starting torque doubles and the locked-rotor current decreases from 100 A to 90 A. The motor develops its breakdown torque at a speed N_d of 500 r/min, compared to the original breakdown speed of 800 r/min.

If we again double the rotor resistance so that it becomes 5 R, the locked-rotor torque attains a maximum value of 250 N·m for a corresponding current of 70 A (Fig. 13.18c).

A further increase in rotor resistance decreases both the locked-rotor torque and locked-rotor current. For example, if the rotor resistance is increased 25 times (25 R), the locked-rotor current drops to 20 A, but the motor develops the same

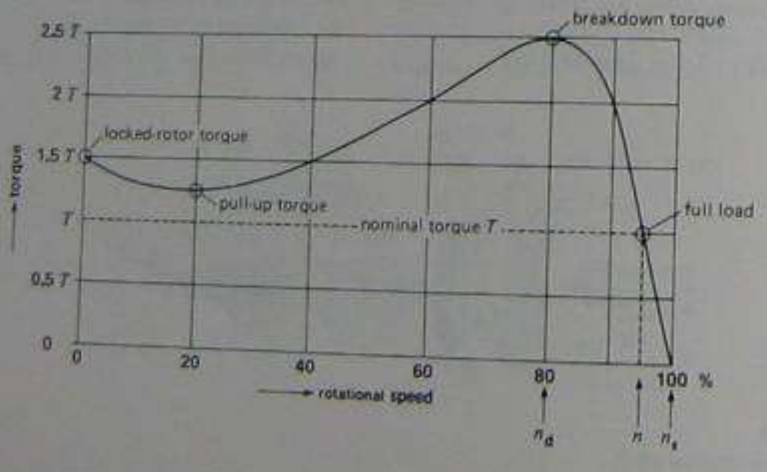


Figure 13.17
Typical torque-speed curve of a 3-phase squirrel-cage induction motor.

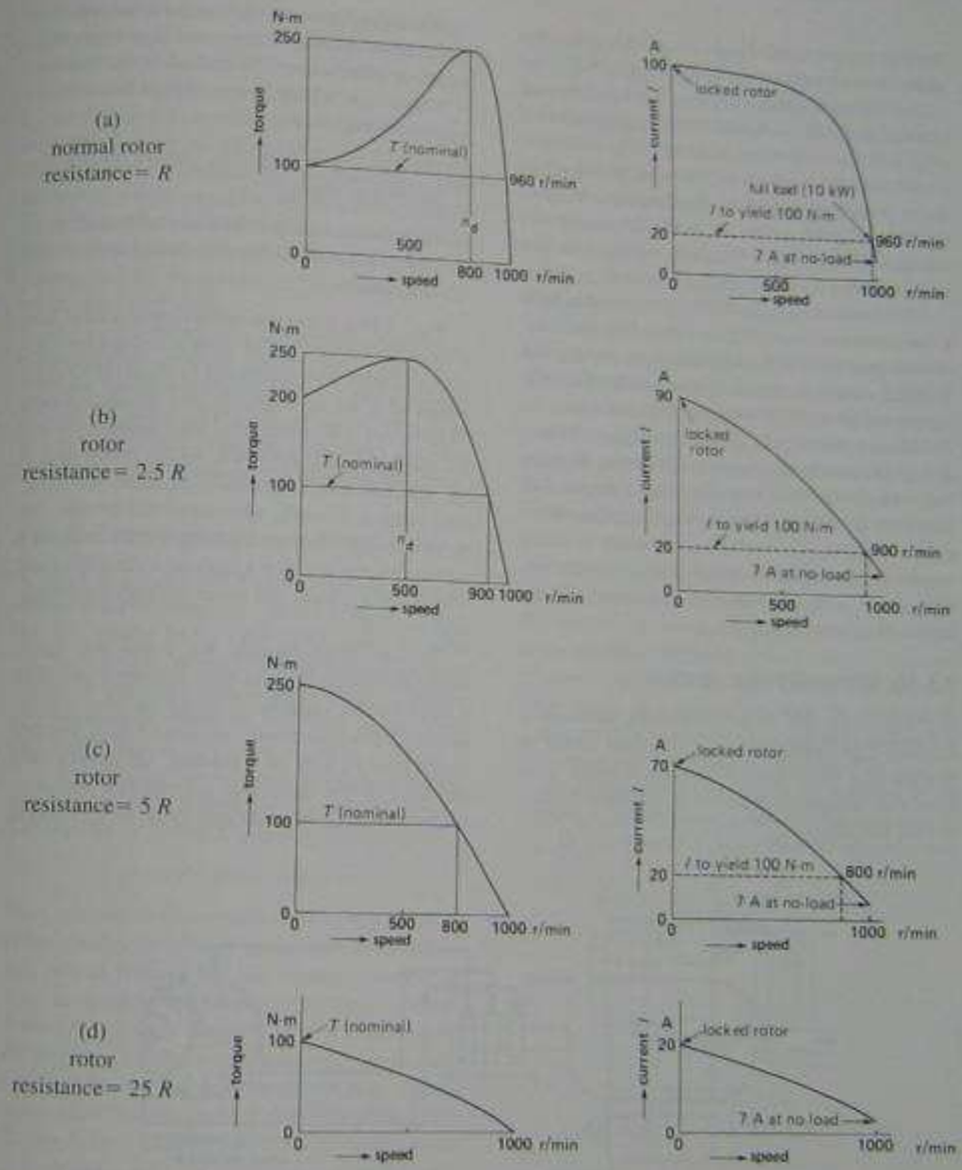


Figure 13.18
Rotor resistance affects the motor characteristics.

Learning Outcome 1.5

Calculate the rotor frequency given a value of supply frequency and slip

13.7 Acceleration of the rotor—slip

As soon as the rotor is released, it rapidly accelerates in the direction of the rotating field. As it picks up speed, the relative velocity of the field with respect to the rotor diminishes progressively. This causes both the value and the frequency of the induced voltage to decrease because the rotor bars are cut more slowly. The rotor current, very large at first, decreases rapidly as the motor picks up speed.

The speed will continue to increase, but it will never catch up with the revolving field. In effect, if the rotor *did* turn at the same speed as the field (synchronous speed), the flux would no longer cut the rotor bars and the induced voltage and current would fall to zero. Under these conditions the force acting on the rotor bars would also become zero and the friction and windage would immediately cause the rotor to slow down.

The rotor speed is always slightly less than synchronous speed so as to produce a current in the rotor bars sufficiently large to overcome the braking torque. At no-load the percent difference in speed between the rotor and field (called *slip*), is small: usually less than 0.1% of synchronous speed.

13.8 Motor under load

Suppose the motor is initially running at no-load. If we apply a mechanical load to the shaft, the motor will begin to slow down and the revolving field will cut the rotor bars at a higher and higher rate. The induced voltage and the resulting current in the bars will increase progressively, producing a greater and greater motor torque. The question is, for how long can this go on? Will the speed continue to drop until the motor comes to a halt?

No, the motor and the mechanical load will reach a state of equilibrium when the motor torque is exactly equal to the load torque. When this state is reached, the speed will cease to drop and the motor will turn at a constant rate. It is very important to understand that a motor only turns at constant speed when its torque is *exactly* equal to the torque exerted by the mechanical load. The moment this state of equilibrium is upset, the motor speed will start to change (Section 3.11).

Under normal loads, induction motors run very close to synchronous speed. Thus, at full-load, the slip for large motors (1000 kW and more) rarely exceeds 0.5% of synchronous speed, and for small machines (10 kW and less), it seldom exceeds 5%. That is why induction motors are considered to be constant speed machines. However, because they never actually turn at synchronous speed, they are sometimes called *asynchronous* machines.

13.9 Slip and slip speed

The slip s of an induction motor is the difference between the synchronous speed and the rotor speed, expressed as a percent (or per-unit) of synchronous speed. The per-unit slip is given by the equation

$$s = \frac{n_s - n}{n_s} \qquad (13.2)$$

where

s = slip
n_s = synchronous speed [r/min]
n = rotor speed [r/min]

The slip is practically zero at no-load and is equal to 1 (or 100%) when the rotor is locked.

Example 13-2

A 0.5 hp, 6-pole induction motor is excited by a 3-phase, 60 Hz source. If the full-load speed is 1140 r/min, calculate the slip.

Solution

The synchronous speed of the motor is

$$n_s = 120f/p = 120 \times 60/6 \qquad (13.1)$$
$$= 1200 \text{ r/min}$$

The difference between the synchronous speed of the revolving flux and rotor speed is the slip speed:

$$n_s - n = 1200 - 1140 = 60 \text{ r/min}$$

The slip is

$$s = (n_s - n)/n_s = 60/1200 \quad (13.2)$$
$$= 0.05 \text{ or } 5\%$$

13.10 Voltage and frequency induced in the rotor

The voltage and frequency induced in the rotor both depend upon the slip. They are given by the following equations:

$$f_2 = sf \quad (13.3)$$
$$E_2 = sE_{oc} \text{ (approx.)} \quad (13.4)$$

where

f_2 = frequency of the voltage and current in the rotor [Hz]

f = frequency of the source connected to the stator [Hz]

s = slip

E_2 = voltage induced in the rotor at slip s

E_{oc} = open-circuit voltage induced in the rotor when at rest [V]

In a cage motor, the open-circuit voltage E_{oc} is the voltage that *would* be induced in the rotor bars if the bars were disconnected from the end-rings. In the case of a wound-rotor motor the open-circuit voltage is $1/\sqrt{3}$ times the voltage between the open-circuit slip-rings.

It should be noted that Eq. 13.3 *always* holds true, but Eq. 13.4 is valid only if the revolving flux (expressed in webers) remains absolutely constant. However, between zero and full-load the actual value of E_2 is only slightly less than the value given by the equation.

Example 13-3

The 6-pole wound-rotor induction motor of Example 13-2 is excited by a 3-phase 60 Hz source. Calculate the frequency of the rotor current under the following conditions:

a. At standstill
b. Motor turning at 500 r/min in the same direction as the revolving field
c. Motor turning at 500 r/min in the opposite direction to the revolving field
d. Motor turning at 2000 r/min in the same direction as the revolving field

Solution

From Example 13-2, the synchronous speed of the motor is 1200 r/min.

a. At standstill the motor speed $n = 0$. Consequently, the slip is

$$s = (n_s - n)/n_s = (1200 - 0)/1200 = 1$$

The frequency of the induced voltage (and of the induced current) is

$$f_2 = sf = 1 \times 60 = 60 \text{ Hz}$$

b. When the motor turns in the same direction as the field, the motor speed n is positive. The slip is

$$s = (n_s - n)/n_s = (1200 - 500)/1200$$
$$= 700/1200 = 0.583$$

The frequency of the induced voltage (and of the rotor current) is

$$f_2 = sf = 0.583 \times 60 = 35 \text{ Hz}$$

c. When the motor turns in the opposite direction to the field, the motor speed is *negative*; thus, $n = -500$. The slip is

$$s = (n_s - n)/n_s$$
$$= [1200 - (-500)]/1200$$
$$= (1200 + 500)/1200 = 1700/1200$$
$$= 1.417$$

A slip greater than 1 implies that the motor is operating as a brake.

The frequency of the induced voltage and rotor current is

$$f_2 = sf = 1.417 \times 60 = 85 \text{ Hz}$$

d. The motor speed is positive because the rotor turns in the same direction as the field: $n = +2000$. The slip is

$$s = (n_s - n)/n_s$$
$$= (1200 - 2000)/1200$$
$$= -800/1200 = -0.667$$

A negative slip implies that the motor is actually operating as a generator.

The frequency of the induced voltage and rotor current is

$$f_2 = sf = -0.667 \times 60 = -40 \text{ Hz}$$

A negative frequency means that the phase sequence of the voltages induced in the rotor windings is reversed. Thus, if the phase sequence of the rotor voltages is A-B-C when the frequency is positive, the phase sequence is A-C-B when the frequency is negative. As far as a frequency meter is concerned, a negative frequency gives the same reading as a positive frequency. Consequently, we can say that the frequency is simply 40 Hz.

13.11 Characteristics of squirrel-cage induction motors

Table 13A lists the typical properties of squirrel-cage induction motors in the power range between 1 kW and 20 000 kW. Note that the current and torque are expressed in per-unit values. The base current is the full-load current and all other currents are compared to it. Similarly, the base torque is the full-load torque and all other torques are compared to it. Finally, the base speed is the synchronous speed of the motor. The following explanations will clarify the meaning of the values given in the table.

1. Motor at no-load. When the motor runs at no load, the stator current lies between 0.5 and 0.3 pu (of full-load current). The no-load current is similar to the exciting current in a transformer. Thus, it is composed of a magnetizing component that creates the revolving flux Φ_m and a small active component that supplies the windage and friction losses in the rotor plus the iron losses in the stator. The flux Φ_m links both the stator and the rotor; consequently it is similar to the mutual flux in a transformer (Fig. 13.13).

Considerable reactive power is needed to create the revolving field and, in order to keep it within acceptable limits, the air gap is made as short as mechanical tolerances will permit. The power factor at no-load is therefore low; it ranges from 0.2 (or 20%) for small machines to 0.05 for large machines. The efficiency is zero because the output power is zero.

2. Motor under load. When the motor is under load, the current in the rotor produces a mmf which tends to change the mutual flux Φ_m. This sets up an opposing current flow in the stator. The opposing mmfs of the rotor and stator are very similar to the opposing mmfs of the secondary and primary in a transformer. As a result, leakage fluxes Φ_{f1} and Φ_{f2} are created, in addition to the mutual flux Φ_m (Fig. 13.14). The total reactive power needed to produce these three fluxes is slightly greater than when the motor is operating at no-load. However, the active power (kW) absorbed by the motor increases in almost direct proportion to the mechanical load. It follows that the power factor

TABLE 13A TYPICAL CHARACTERISTICS OF SQUIRREL-CAGE INDUCTION MOTORS

Loading	Current (per-unit)		Torque (per-unit)		Slip (per-unit)		Efficiency		Power factor	
Motor size →	Small*	Big*	Small	Big	Small	Big	Small	Big	Small	Big
Full-load	1	1	1	1	0.03	0.004	0.7 to 0.9	0.96 to 0.98	0.8 to 0.85	0.87 to 0.9
No-load	0.5	0.3	0	0	~0	~0	0	0	0.2	0.05
Locked rotor	5 to 6	4 to 6	1.5 to 3	0.5 to 1	1	1	0	0	0.4	0.1

*Small means under 11 kW (15 hp); big means over 1120 kW (1500 hp) and up to 25 000 hp.

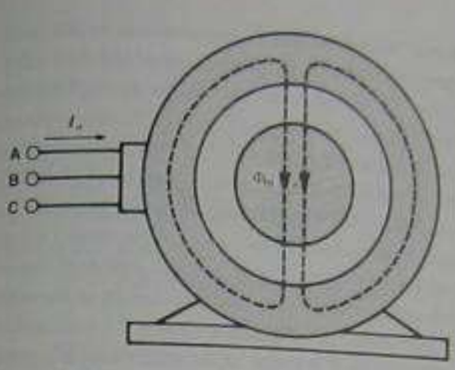


Figure 13.13
At no-load the flux in the motor is mainly the mutual flux Φ_m. To create this flux, considerable reactive power is needed.

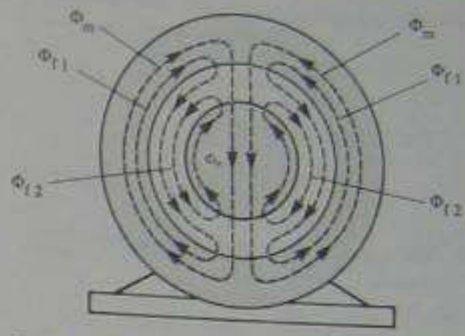


Figure 13.14
At full-load the mutual flux decreases, but stator and rotor leakage fluxes are created. The reactive power needed is slightly greater than in Fig. 13.13.

of the motor improves dramatically as the mechanical load increases. At full-load it ranges from 0.80 for small machines to 0.90 for large machines. The efficiency at full-load is particularly high; it can attain 98% for very large machines.

*3. **Locked-rotor characteristics.*** The locked-rotor current is 5 to 6 times the full-load current, making the I^2R losses 25 to 36 times higher than normal. The rotor must therefore never remain locked for more than a few seconds.

Although the mechanical power at standstill is zero, the motor develops a strong torque. The power factor is low because considerable reactive power is needed to produce the leakage flux in the rotor and stator windings. These leakage fluxes are much larger than in a transformer because the stator and the rotor windings are not as tightly coupled (see Section 10.2).

13.12 Estimating the currents in an induction motor

The full-load current of a 3-phase induction motor may be calculated by means of the following approximate equation:

$$I = 600\,P_h/E \qquad (13.5)$$

where

I = full-load current [A]
P_h = output power [horsepower]
E = rated line voltage (V)
600 = empirical constant

Recalling that the starting current is 5 to 6 pu and that the no-load current lies between 0.5 and 0.3 pu, we can readily estimate the value of these currents for any induction motor.

Example 13-4

a. Calculate the approximate full-load current, locked-rotor current, and no-load current of a 3-phase induction motor having a rating of 500 hp, 2300 V.
b. Estimate the apparent power drawn under locked-rotor conditions.
c. State the nominal rating of this motor, expressed in kilowatts.

Solution

a. The full-load current is

$$\begin{aligned} I &= 600\,P_h/E \qquad (13.5) \\ &= 600 \times 500/2300 \\ &= 130\ \text{A (approx.)} \end{aligned}$$

The no-load current is

$$I_o = 0.3I = 0.3 \times 130$$
$$= 39 \text{ A (approx.)}$$

The starting current is

$$I_{LR} = 6I = 6 \times 130$$
$$= 780 \text{ A (approx.)}$$

b. The apparent power under locked-rotor conditions is

$$S = \sqrt{3}\,EI$$
$$= \sqrt{3} \times 2300 \times 780 \qquad (8.9)$$
$$= 3100 \text{ kVA (approx.)}$$

c. When the power of a motor is expressed in kilowatts, it always relates to the mechanical output and *not* to the electrical input. The nominal rating of this motor expressed in SI units is, therefore,

$$P = 500/1.34$$
$$= 373 \text{ kW (see Power conversion chart in Appendix AX0)}$$

13.13 Active power flow

Voltages, currents, and phasor diagrams enable us to understand the detailed behavior of an induction motor. However, it is easier to see how electrical energy is converted into mechanical energy by following the active power as it flows through the machine. Thus, referring to Fig. 13.15, active power P_e flows from the line into the 3-phase stator. Due to the stator copper losses, a portion P_{js} is dissipated as heat in the windings. Another portion P_f is dissipated as heat in the stator core, owing to the iron losses. The remaining active power P_r is carried across the air gap and transferred to the rotor by electromagnetic induction.

Due to the I^2R losses in the rotor, a third portion P_{jr} is dissipated as heat, and the remainder is finally available in the form of mechanical power P_m. By subtracting a small fourth portion P_v, representing windage and bearing-friction losses, we finally obtain P_L, the mechanical power available at the shaft to drive the load.

The power flow diagram of Fig. 13.15 enables us to identify and to calculate three important properties of the induction motor: (1) its *efficiency*, (2) its *power*, and (3) its *torque*.

1. Efficiency. By definition, the efficiency of a motor is the ratio of the output power to the input power:

$$\text{efficiency } (\eta) = P_L/P_e \qquad (13.6)$$

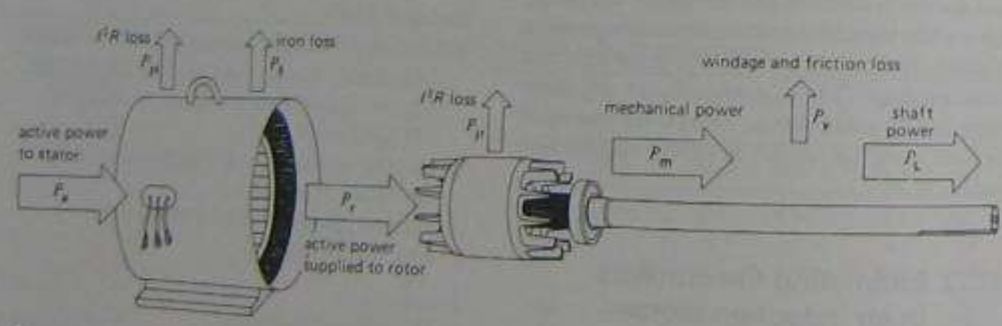


Figure 13.15
Active power flow in a 3-phase induction motor.

2. I^2R losses in the rotor. It can be shown* that the rotor I^2R losses P_{jr} are related to the rotor input power P_r by the equation

$$P_{jr} = sP_r \tag{13.7}$$

where

P_{jr} = rotor I^2R losses [W]

s = slip

P_r = power transmitted to the rotor [W]

Equation 13.7 shows that as the slip increases, the rotor I^2R losses consume a larger and larger proportion of the power P_r transmitted across the air gap to the rotor. A rotor turning at half synchronous speed ($s = 0.5$) dissipates in the form of heat 50 percent of the active power it receives. When the rotor is locked ($s = 1$), all the power transmitted to the rotor is dissipated as heat.

3. Mechanical power. The mechanical power P_m developed by the motor is equal to the power transmitted to the rotor minus its I^2R losses. Thus,

$$P_m = P_r - P_{jr}$$
$$= P_r - sP_r \tag{13.7}$$

whence

$$P_m = (1 - s)P_r \tag{13.8}$$

The actual mechanical power available to drive the load is slightly less than P_m, due to the power needed to overcome the windage and friction losses. In most calculations we can neglect this small loss.

4. Motor torque. The torque T_m developed by the motor at *any* speed is given by

$$T_m = \frac{9.55\,P_m}{n} \tag{3.5}$$

$$= \frac{9.55\,(1-s)\,P_r}{n_s(1-s)} = 9.55\,P_r/n_s$$

therefore,

$$T_m = 9.55\,P_r/n_s \tag{13.9}$$

where

T_m = torque developed by the motor at *any* speed [N·m]

P_r = power transmitted to the rotor [W]

n_s = synchronous speed [r/min]

9.55 = multiplier to take care of units [exact value: $60/2\pi$]

The actual torque T_L available at the shaft is slightly less than T_m, due to the torque required to overcome the windage and friction losses. However, in most calculations we can neglect this small difference.

Equation 13.9 shows that the torque is directly proportional to the active power transmitted to the rotor. Thus, to develop a high locked-rotor torque, the rotor must absorb a large amount of active power. The latter is dissipated in the form of heat, consequently, the temperature of the rotor rises very rapidly.

Example 13-5

A 3-phase induction motor having a synchronous speed of 1200 r/min draws 80 kW from a 3-phase

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* $\begin{bmatrix}\text{mechanical}\\ \text{power output}\\ \text{of rotor}\end{bmatrix} = \begin{bmatrix}\text{electromagnetic}\\ \text{power transferred}\\ \text{to rotor}\end{bmatrix} - \begin{bmatrix}\text{electrical}\\ \text{losses}\\ \text{in rotor}\end{bmatrix}$

$$P_m = P_r - P_{jr} \tag{i}$$

but from Eq. 3.5

$$P_m = \frac{\text{rotor speed} \times \text{mechanical torque}}{9.55}$$

Hence,

$$P_m = \frac{nT_m}{9.55} \tag{ii}$$

Also from Eq. 3.5 we can write

$$P_r = \frac{\text{speed of flux} \times \text{electromagnetic torque}}{9.55}$$

$$P_r = \frac{n_s T_{mag}}{9.55} \tag{iii}$$

but the mechanical torque T_m must equal the electromagnetic torque T_{mag}.

Thus

$$T_m = T_{mag} \tag{iv}$$

Substituting (ii), (iii), and (iv) in (i), we find

$$P_{jr} = sP_r$$

feeder. The copper losses and iron losses in the stator amount to 5 kW. If the motor runs at 1152 r/min, calculate the following:

a. The active power transmitted to the rotor
b. The rotor I^2R losses
c. The mechanical power developed
d. The mechanical power delivered to the load, knowing that the windage and friction losses are equal to 2 kW
e. The efficiency of the motor

Solution

a. Active power to the rotor is

$$P_r = P_e - P_{js} - P_f$$
$$= 80 - 5 = 75 \text{ kW}$$

b. The slip is

$$s = (n_s - n)/n_s$$
$$= (1200 - 1152)/1200$$
$$= 48/1200 = 0.04$$

Rotor I^2R losses are

$$P_{jr} = sP_r = 0.04 \times 75 = 3 \text{ kW}$$

c. The mechanical power developed is

$$P_m = P_r - I^2R \text{ losses in rotor}$$
$$= 75 - 3 = 72 \text{ kW}$$

d. The mechanical power P_L delivered to the load is slightly less than P_m, due to the friction and windage losses.

$$P_L = P_m - P_v = 72 - 2 = 70 \text{ kW}$$

e. The efficiency is

$$\eta = P_L/P_e = 70/80$$
$$= 0.875 \text{ or } 87.5\%$$

Example 13-6

A 3-phase, 8-pole squirrel-cage induction motor, connected to a 60 Hz line, possesses a synchronous speed of 900 r/min. The motor absorbs 40 kW, and the copper and iron losses in the stator amount to 5 kW and 1 kW, respectively. Calculate the torque developed by the motor.

Solution

The power transmitted across the air gap to the rotor is

$$P_r = P_e - P_{js} - P_f$$
$$= 40 - 5 - 1 = 34 \text{ kW}$$

$$T_m = 9.55\ P_r/n_s \qquad (13.9)$$
$$= 9.55 \times 34\,000/900$$
$$= 361 \text{ N·m}$$

Note that the solution to this problem (the torque) is independent of the speed of rotation. The motor could be at a standstill or running at full speed, but as long as the power P_r transmitted to the rotor is equal to 34 kW, the motor develops a torque of 361 N·m.

Learning Outcome 1.3 Continue

Winding Calculation

&

Learning Outcome 1..7

Perform calculations using the relationship between air gap power, losses and net torque

Figure 13.22a
The 24 coils are grouped two-by-two to make 12 groups.

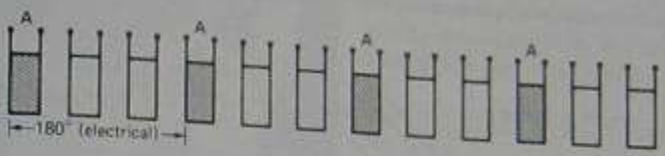


Figure 13.22b
The four groups of phase A are selected so as to be evenly spaced from each other.

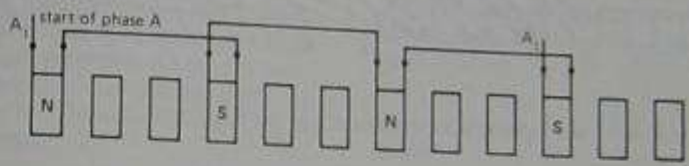


Figure 13.22c
The groups of phase A are connected in series to create alternate N-S poles.

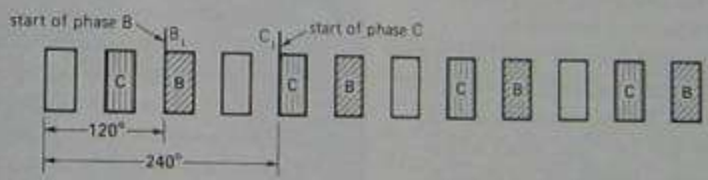


Figure 13.22d
The start of phases B and C begins 120° and 240°, respectively, after the start of phase A.

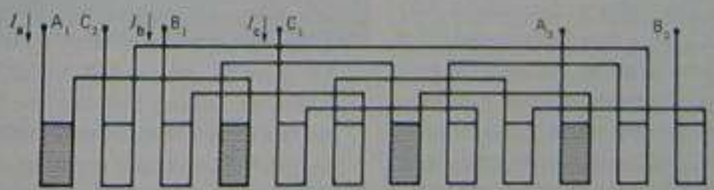


Figure 13.22e
When all phase groups are connected, only six leads remain.

coil side set in each slot. If the windings are now laid down so that all the other coil sides fall into the slots, we obtain the classical appearance of a 3-phase lap winding having two coil sides per slot (Fig. 13.21b). The coils are connected together to create three identical windings, one for each phase. Each winding consists of a number of groups equal to the number of poles. The groups of each phase are symmetrically distributed around the circumference of the stator. The following examples show how this is done.

Example 13-8

The stator of a 3-phase, 10-pole induction motor possesses 120 slots. If a lap winding is used, calculate the following:

a. The total number of coils
b. The number of coils per phase
c. The number of coils per group
d. The pole pitch
e. The coil pitch (expressed as a percentage of the pole pitch), if the coil width extends from slot 1 to slot 11

Solution

a. A 120-slot stator requires 120 coils.
b. Coils per phase = $120 \div 3 = 40$.

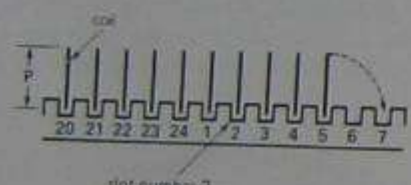


Figure 13.21a
Coils held upright in 24 stator slots.

Figure 13.21b
Coils laid down to make a typical lap winding.

c. Number of groups per phase = number of poles = 10
Coils per group = $40 \div 10 = 4$.
d. The pole pitch corresponds to

$$\text{pole pitch} = \text{slots/poles} = 120/10 = 12 \text{ slots}$$

One pole pitch extends therefore from slot 1 (say) to slot 13.
e. The coil pitch covers 10 slots (slot 1 to slot 11). The percent coil pitch = $10/12 = 83.3\%$.

The next example shows in greater detail how the coils are interconnected in a typical 3-phase stator winding.

Example 13-9

A stator having 24 slots has to be wound with a 3-phase, 4-pole winding. Determine the following:
1. The connections between the coils
2. The connections between the phases

Solution

The 3-phase winding has 24 coils. Assume that they are standing upright, with one coil side in each slot (Fig. 13.22). We will first determine the coil distribution for phase A and then proceed with the connections for that phase. Similar connections will then be made for phases B and C. Here is the line of reasoning:

a. The revolving field creates 4 poles; the motor therefore has 4 groups per phase, or $4 \times 3 = 12$ phase groups in all. Each rectangle in Fig. 13.22a represents one group. Because the stator contains 24 coils, each group consists of $24/12 = 2$ consecutive coils.
b. The groups (poles) of each phase must be uniformly spaced around the stator. The group distribution for phase A is shown in Fig. 13.22b. Each shaded rectangle represents two upright coils connected in series, producing the two terminals shown. Note that the mechanical distance between two successive groups always corresponds to an *electrical* phase angle of 180°.
c. Successive groups of phase A must have opposite magnetic polarities. Consequently, the four

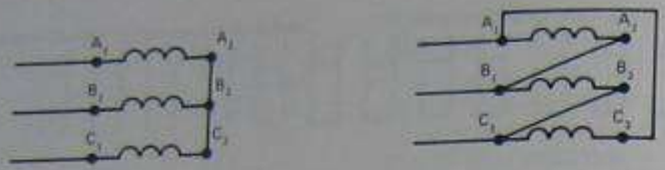


Figure 13.22f
The phase may be connected in wye or in delta, and three leads are brought out to the terminal box.

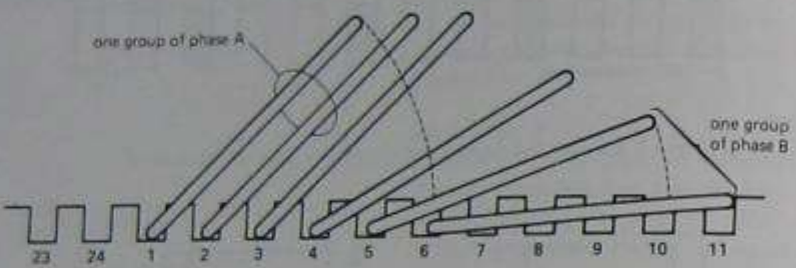


Figure 13.23
The pole pitch is from slot 1 slot 7; the coil pitch from slot 1 to slot 6.

groups of phase A are connected in series to produce successive N-S-N-S poles (Fig. 13.22c). Phase A now has two terminals, a *starting* terminal A_1 and a *finishing* terminal A_2.

d. The phase groups of phases B and C are spaced the same way around the stator. However, the *starting* terminals B_1 and C_1 are respectively located at 120° and 240° (electrical) with respect to the starting terminal A_1 of phase A (Fig. 13.22d).

e. The groups in phases B and C are connected in series in the same way as those of phase A are (Fig. 13.22e). This yields six terminals: A_1A_2, B_1B_2, and C_1C_2. They may be connected either in wye or in delta inside the machine. The resulting 3 wires corresponding to the 3 phases are brought out to the terminal box of the machine (Fig. 13.22f). In practice, the connections are made, not while the coils are upright (as shown) but only after they have been laid down in the slots.

f. Because the pole pitch corresponds to a span of $24/4 = 6$ slots, the coil pitch may be shortened to 5 slots (slot 1 to slot 6). Thus, the first coil of phase A is lodged in the first and sixth slots (Fig. 13.23). All the other coils and connections follow suit according to Fig. 13.22e.

Figs. 13.24a and 13.24b show the coil and stator of a 450 kW (600 hp) induction motor. Fig. 13.25 illustrates the procedure used in winding a smaller 37.5 kW (50 hp) stator.

13.18 Sector motor

Consider a standard 3-phase, 4-pole, wye-connected motor having a synchronous speed of 1800 r/min. Let us cut the stator in half, so that half the winding is removed and only two complete N and S poles are left (per phase). Next, let us connect the three phases in wye, without making any other changes to the existing coil connections. Finally, we mount the original rotor above this *sector stator*, leaving a small air gap (Fig. 13.26).

Figure 13.24a
Stator of a 3-phase, 450 kW, 1180 r/min, 575 V, 60 Hz induction motor. The lap winding is composed of 108 preformed coils having a pitch from slots 1 to 15. One coil side falls into the bottom of a slot and the other at the top. Rotor diameter: 500 mm; axial length: 460 mm. *(Courtesy of Services Électromécaniques Roberge)*

Figure 13.24b
Close-up view of the preformed coil in Fig. 13.24a.

If we connect the stator terminals to a 3-phase, 60 Hz source, the rotor will again turn at close to 1800 r/min. To prevent saturation, the voltage should be reduced to half its original value because the stator winding now has only one-half the original number of turns. Under these conditions, this remarkable truncated *sector motor* still develops about 20 percent of its original rated power.

The sector motor produces a *revolving* field that moves at the same peripheral speed as the flux in the original 3-phase motor. However, instead of making a complete turn, the field simply travels continuously from one end of the stator to the other.

13.19 Linear induction motor

It is obvious that the sector stator could be laid out flat, without affecting the shape or speed of the magnetic field. Such a flat stator produces a field that moves at constant speed, in a straight line. Using the same reasoning as in Section 13.5, we can prove that the flux travels at a linear synchronous speed given by

$$v_s = 2\,wf \qquad (13.10)$$

where

v_s = linear synchronous speed [m/s]
w = width of one pole-pitch [m]
f = frequency [Hz]

Note that the linear speed does not depend upon the number of poles but only on the pole-pitch. Thus, it is possible for a 2-pole linear stator to create a field moving at the same speed as that of a 6-pole linear stator (say), provided they have the same pole-pitch.

If a flat squirrel-cage winding is brought near the flat stator, the travelling field drags the squirrel cage along with it (Section 13.2). In practice, we generally use a simple aluminum or copper plate as a rotor (Fig. 13.27). Furthermore, to increase the power and to reduce the reluctance of the magnetic path, two flat stators are usually mounted, face-to-face, on opposite sides of the aluminum plate. The combination is called a *linear induction motor*. The direction of the motor can be reversed by interchanging any two stator leads.

In many practical applications, the rotor is stationary while the stator moves. For example, in some high-speed trains, the rotor is composed of a

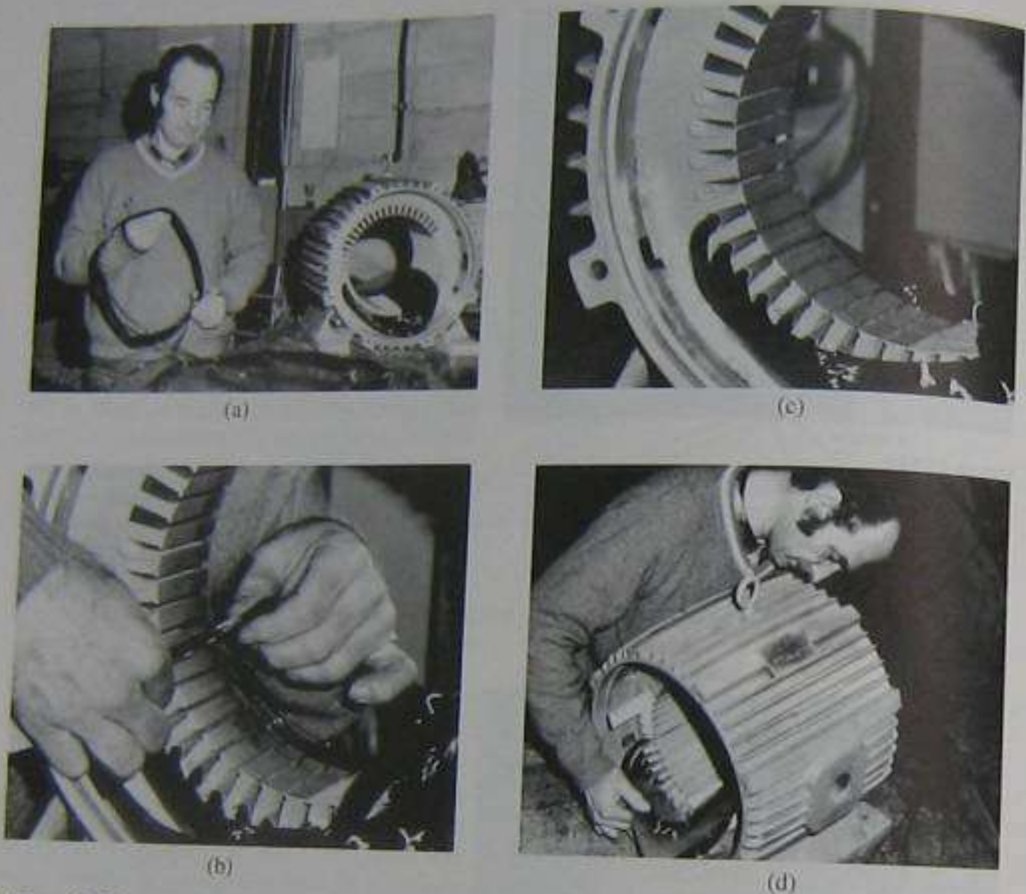


Figure 13.25

Stator winding of a 3-phase, 50 hp, 575 V, 60 Hz, 1764 r/min induction motor. The stator possesses 48 slots carrying 48 coils connected in wye.

a. Each coil is composed of 5 turns of five No. 15 copper wires connected in parallel. The wires are covered with a high-temperature polyimide insulation. Five No. 15 wires in parallel is equivalent to one No. 8 wire.
b. One coil side is threaded into slot 1 (say) and the other side goes into slot 12. The coil pitch is, therefore, from 1 to 12.
c. Each coil side fills half a slot and is covered with a paper spacer so that it does not touch the second coil side placed in the same slot. Starting from the top, the photograph shows 3 empty and uninsulated slots and 4 empty slots insulated with a composition paper liner. The remaining 10 slots each carry one coil side.
d. A varnished cambric cloth, cut in the shape of a triangle, provides extra insulation between adjacent phase groups.

(Courtesy of Services Électromécaniques Roberge)

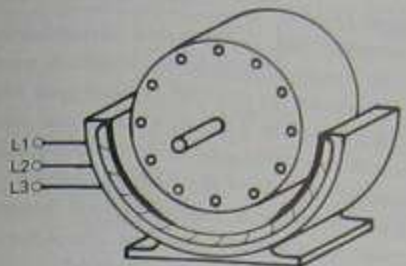


Figure 13.26
Two-pole sector induction motor.

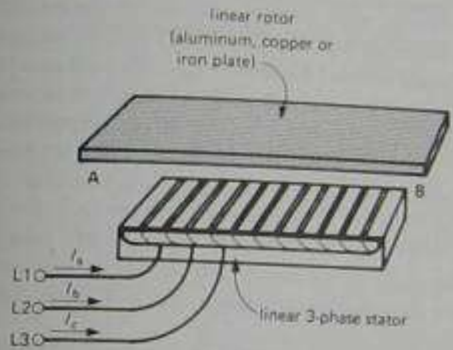


Figure 13.27
Components of a 3-phase linear induction motor.

thick aluminum plate fixed to the ground and extending over the full length of the track. The linear stator is bolted to the undercarriage of the train and straddles the plate. Train speed is varied by changing the frequency applied to the stator (Fig. 13.31).

Example 13-10

The stator of a linear induction motor is excited from a 75 Hz electronic source. If the distance between consecutive phase groups of phase A is 300 mm, calculate the linear speed of the magnetic field.

Solution

The pole pitch is 300 mm. Consequently,

$$v_s = 2\,wf \qquad (13.10)$$
$$= 2 \times 0.3 \times 75$$
$$= 45 \text{ m/s or } 162 \text{ km/h}$$

13.20 Traveling waves

We are sometimes left with the impression that when the flux reaches the end of a linear stator, there must be a delay before it returns to restart once more at the beginning. This is not the case. The linear motor produces a traveling wave of flux which moves continuously and smoothly from one end of the stator to the other. Figure 13.28 shows how the flux moves from left to right in a 2-pole linear motor. The flux cuts off sharply at extremities A, B of the stator. However, as fast as a N or S pole disappears at the right, it builds up again at the left.

13.21 Properties of a linear induction motor

The properties of a linear induction motor are almost identical to those of a standard rotating machine. Consequently, the equations for slip, thrust, power, etc., are also similar.

1. Slip. Slip is defined by

$$s = (v_s - v)/v_s \qquad (13.11)$$

where

s = slip
v_s = synchronous linear speed [m/s]
v = speed of rotor (or stator) [m/s]

2. Active power flow. With reference to Fig. 13.15, active power flows through a linear motor in the same way it does through a rotating motor, except that the stator and rotor are flat. Consequently, Eqs. 13.6, 13.7, and 13.8 apply to both types of machines:

$$\eta = P_L/P_e \qquad (13.6)$$
$$P_{jr} = sP_r \qquad (13.7)$$
$$P_m = (1 - s)P_r \qquad (13.8)$$

3. Thrust. The thrust or force developed by a linear induction motor is given by:

$$F = P_r/v_s \qquad (13.12)$$

where

F = thrust [N]
P_r = power transmitted to the rotor [W]
v_s = linear synchronous speed [m/s]

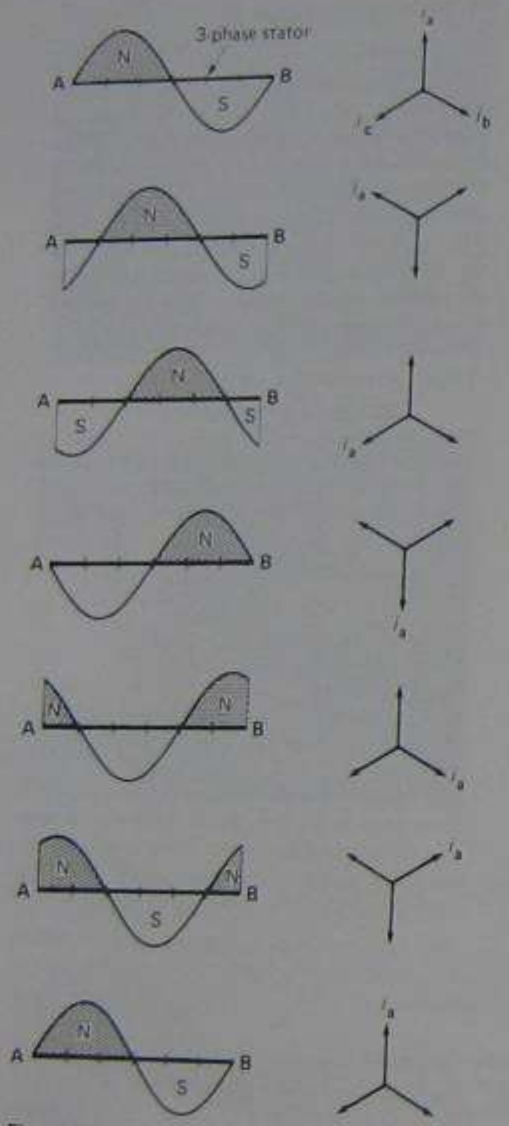


Figure 13.28
Shape of the magnetic field created by a 2-pole, 3-phase linear stator, over one complete cycle. The successive frames are separated by an interval of time equal to 1/6 cycle or 60°.

Example 13-11

An overhead crane in a factory is driven horizontally by means of two linear induction motors whose rotors are the two steel I-beams upon which the crane rolls. The 3-phase, 4-pole linear stators (mounted on opposite sides of the crane and facing the respective webs of the I-beams) have a pole pitch of 8 cm and are driven by a variable frequency electronic source. During a test on one of the motors, the following results were obtained:

stator frequency: 15 Hz
power to stator: 5 kW
copper loss + iron loss in stator: 1 kW
crane speed: 1.8 m/s

Calculate

a. Synchronous speed and slip
b. Power to the rotor
c. I^2R loss in rotor
d. Mechanical power and thrust

Solution

a. Linear synchronous speed

$$\begin{aligned} v_s &= 2\,wf \qquad (13.10)\\ &= 2 \times 0.08 \times 15\\ &= 2.4 \text{ m/s} \end{aligned}$$

The slip is

$$\begin{aligned} s &= (v_s - v)/v_s \qquad (13.11)\\ &= (2.4 - 1.8)/2.4\\ &= 0.25 \end{aligned}$$

b. Power to the rotor is

$$\begin{aligned} P_r &= P_e - P_{js} - P_f \qquad \text{(see Fig. 13.15)}\\ &= 5 - 1\\ &= 4 \text{ kW} \end{aligned}$$

c. I^2R loss in the rotor is

$$\begin{aligned} P_{jr} &= sP_r \qquad (13.7)\\ &= 0.25 \times 4\\ &= 1 \text{ kW} \end{aligned}$$

d. Mechanical power is

$$P_m = P_r - P_{jr} \quad \text{(Fig. 13.15)}$$
$$= 4 - 1$$
$$= 3 \text{ kW}$$

The thrust is

$$F = P_r/v_s \quad (13.12)$$
$$= 4000/2.4$$
$$= 1667 \text{ N} = 1.67 \text{ kN } (\sim 375 \text{ lb})$$

13.22 Magnetic levitation

In Section 13.2 we saw that a moving permanent magnet, sweeping across a conducting ladder, tends to drag the ladder along with the magnet. We will now show that this horizontal tractive force is also accompanied by a *vertical* force, which tends to push the magnet away from the ladder.

Referring to Fig. 13.29, suppose that conductors 1, 2, 3 are three conductors of the stationary ladder. The center of the N pole of the magnet is sweeping across the top of conductor 2. The voltage induced in this conductor is maximum because the flux density is greatest at the center of the pole. If the magnet moves very slowly, the resulting induced current reaches *its* maximum value at virtually the same time. This current, returning by conductors 1 and 3, creates magnetic poles nnn and sss as shown in Fig. 13.29. According to the laws of attraction and repulsion, the front half of the magnet is repelled upward while the rear half is attracted downward. Because the distribution of the nnn and sss poles is symmetrical with respect to the center of the magnet, the vertical forces of attraction and repulsion are equal, and the resulting vertical force is nil. Consequently, there is only a horizontal tractive force.

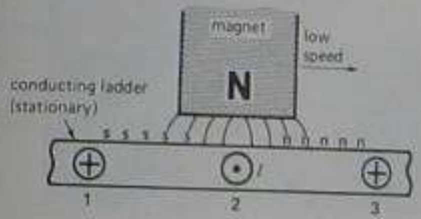


Figure 13.29
Currents and magnetic poles at low speed.

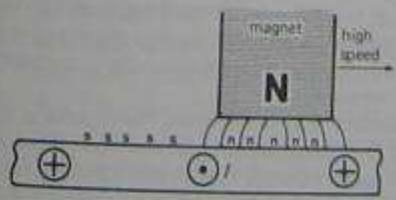


Figure 13.30
Currents and magnetic poles at high speed.

But suppose now that the magnet moves very rapidly. Owing to its inductance, the current in conductor 2 reaches its maximum value a fraction of a second after the voltage has attained *its* maximum. Consequently, by the time the current in conductor 2 is maximum, the center of the magnet is already some distance ahead of the conductor (Fig. 13.30). The current returning by conductors 1 and 3 again creates nnn and sss poles; however, the N pole of the magnet is now directly above an nnn pole, with the result that a strong vertical force tends to push the magnet upward.* This effect is called the principle of *magnetic levitation*.

Magnetic levitation is used in some ultra-high-speed trains that glide on a magnetic cushion rather than on wheels. A powerful electromagnet fixed underneath the train moves above a conducting rail inducing currents in the rail in the same way as in our ladder. The force of levitation is always accompanied by a small horizontal braking force which must, of course, be overcome by the linear motor that propels the train. See Figs. 13.31 and 13.32.

* The current is always delayed (even at low speeds) by an interval of time Δt, which depends upon the L/R time constant of the rotor. This delay is so brief that, at low speeds, the current reaches its maximum at virtually the same time and place as the voltage does. On the other hand, at high speeds, the same delay Δt produces a significant shift in space between the points where the voltage and current reach their respective maximum values.

ELECTROMAGNETISM

2.16 Magnetic field intensity *H* and flux density *B*

Whenever a magnetic flux ϕ exists in a body or component, it is due to the presence of a magnetic field *intensity* H, given by

$$H = U/l \qquad (2.18)$$

where

H = magnetic field intensity [A/m]
U = magnetomotive force acting on the component [A] (or ampere turn)
l = length of the component [m]

The resulting magnetic flux density is given by

$$B = \phi/A \qquad (2.19)$$

where

B = flux density [T]
ϕ = flux in the component [Wb]
A = cross section of the component [m^2]

There is a definite relationship between the flux density (B) and the magnetic field intensity (H) of any material. This relationship is usually expressed graphically by the B-H curve of the material.

2.17 *B-H* curve of vacuum

In vacuum, the magnetic flux density B is directly proportional to the magnetic field intensity H, and is expressed by the equation

$$B = \mu_o H \qquad (2.20)$$

where

B = flux density [T]
H = magnetic field intensity [A/m]
μ_o = magnetic constant [$= 4\pi \times 10^{-7}$]*

* Also called the permeability of vacuum. The complete expression for μ_o is $4\pi \times 10^{-7}$ henry/meter.

In the SI, the magnetic constant is fixed, by definition. It has a numerical value of $4\pi \times 10^{-7}$ or approximately 1/800 000. This enables us to write Eq. 2-20 in the approximate form:

$$H = 800\,000\, B \qquad (2.21)$$

The B-H curve of vacuum is a straight line. A vacuum never saturates, no matter how great the flux density may be (Fig. 2.25). The curve shows that a magnetic field intensity of 800 A/m produces a flux density of 1 millitesla.

Nonmagnetic materials such as copper, paper, rubber, and air have B-H curves almost identical to that of vacuum.

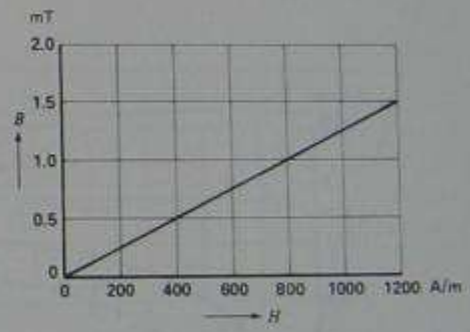


Figure 2.25
B-H curve of vacuum and of nonmagnetic materials.

2.18 *B-H* curve of a magnetic material

The flux density in a magnetic material also depends upon the magnetic field intensity to which it is subjected. Its value is given by

$$B = \mu_o \mu_r H \qquad (2.22)$$

where B, μ_o, and H have the same significance as before, and μ_r is the relative permeability of the material.

The value of μ_r is not constant but varies with the flux density in the material. Consequently, the relationship between B and H is not linear, and this makes Eq. 2.22 rather impractical to use. We prefer to show the relationship by means of a B-H saturation curve. Thus, Fig. 2.26 shows typical saturation curves of three materials commonly used in electrical machines: silicon iron, cast iron, and cast steel. The curves show that a magnetic field intensity of 2000 A/m produces a flux density of 1.4 T in cast steel but only 0.5 T in cast iron.

2.19 Determining the relative permeability

The *relative permeability* μ_r of a material is the ratio of the flux density in the material to the flux density that would be produced in vacuum, under the same magnetic field intensity H.

Given the saturation curve of a magnetic material, it is easy to calculate the relative permeability using the approximate equation

$$\mu_r \approx 800\,000\ B/H \qquad (2.23)$$

where

B = flux density in the magnetic material [T]

H = corresponding magnetic field intensity [A/m]

Example 2-7

Determine the permeability of silicon iron (1%) at a flux density of 1.4 T.

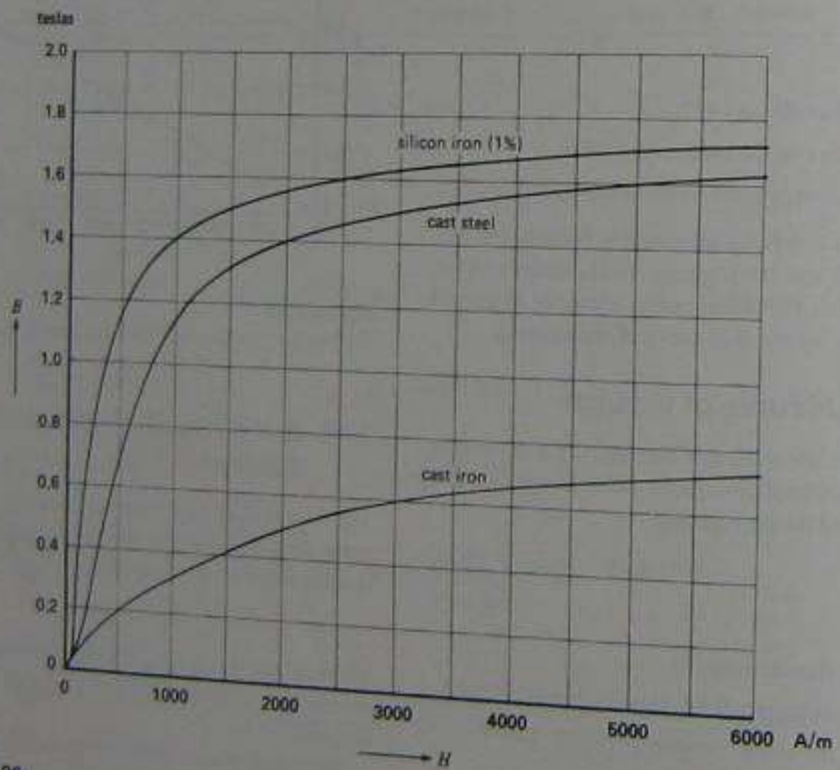


Figure 2.26
B-H saturation curves of three magnetic materials.

Solution

Referring to the saturation curve (Fig. 2.26), we see that a flux density of 1.4 T requires a magnetic field intensity of 1000 A/m. Consequently,

$$\mu_r = 800\,000\ B/H$$
$$= 800\,000 \times 1.4/1000 = 1120$$

At this flux density, silicon iron is 1120 times more permeable than vacuum (or air).

Fig. 2.27 shows the saturation curves of a broad range of materials from vacuum to Permalloy®, one of the most permeable magnetic materials known. Note that as the magnetic field intensity increases, the magnetic materials saturate more and more and eventually all the *B-H* curves follow the *B-H* curve of vacuum.

2.20 Faraday's law of electromagnetic induction

In 1831, while pursuing his experiments, Michael Faraday made one of the most important discoveries in electromagnetism. Now known as **Faraday's law of electromagnetic induction**, it revealed a fundamental relationship between the voltage and flux in a circuit. Faraday's law states:

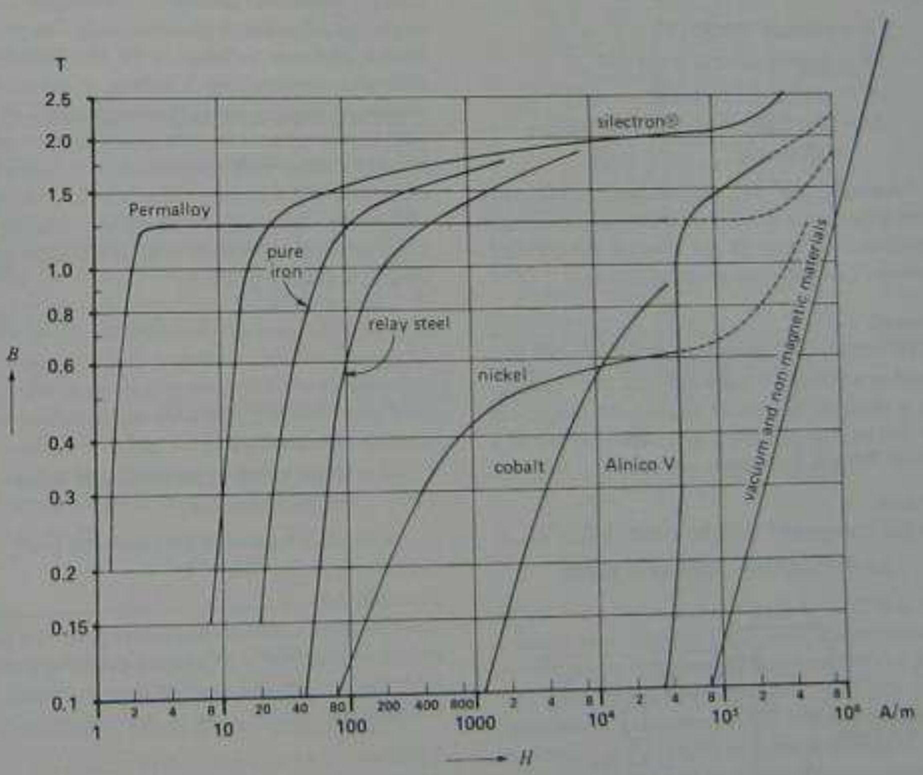


Figure 2.27
Saturation curves of magnetic and nonmagnetic materials. Note that all curves become asymptotic to the *B-H* curve of vacuum where *H* is high.

1. **If the flux linking a loop (or turn) varies as a function of time, a voltage is induced between its terminals.**
2. **The value of the induced voltage is proportional to the rate of change of flux.**

By definition, and according to the SI system of units, when the flux inside a loop varies at the rate of 1 weber per second, a voltage of 1 V is induced between its terminals. Consequently, if the flux varies inside a coil of N turns, the voltage induced is given by

$$E = N\frac{\Delta\Phi}{\Delta t} \qquad (2.24)$$

where

E = induced voltage [V]
N = number of turns in the coil
$\Delta\Phi$ = change of flux inside the coil [Wb]
Δt = time interval during which the flux changes [s]

Faraday's law of electromagnetic induction opened the door to a host of practical applications and established the basis of operation of transformers, generators, and alternating current motors.

Example 2-8

A coil of 2000 turns surrounds a flux of 5 mWb produced by a permanent magnet (Fig. 2.28). The magnet is suddenly withdrawn causing the flux inside the coil to drop uniformly to 2 mWb in 1/10 of a second. What is the voltage induced?

Solution

The flux change is

$$\Delta\Phi = (5\text{ mWb} - 2\text{ mWb}) = 3\text{ mWb}$$

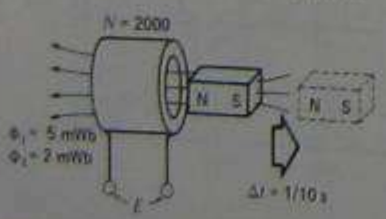


Figure 2.28
Voltage induced by a moving magnet. See Example 2-8.

Because this change takes place uniformly in 1/10 of a second (Δt), the induced voltage is

$$E = N\frac{\Delta\Phi}{\Delta t} = 2000 \times \frac{3}{1000 \times 1/10}$$

$$= 60\text{ V}$$

The induced voltage falls to zero as soon as the flux ceases to change.

2.21 Voltage induced in a conductor

In many motors and generators, the coils move with respect to a flux that is fixed in space. The relative motion produces a change in the flux linking the coils and, consequently, a voltage is induced according to Faraday's law. However, in this special (although common) case, it is easier to calculate the induced voltage with reference to the *conductors*, rather than with reference to the coil itself. In effect, whenever a conductor cuts a magnetic field, a voltage is induced across its terminals. The value of the induced voltage is given by

$$E = Blv \qquad (2.25)$$

where

E = induced voltage [V]
B = flux density [T]
l = active length of the conductor in the magnetic field [m]
v = relative speed of the conductor [m/s]

Example 2-9

The stationary conductors of a large generator have an active length of 2 m and are cut by a field of 0.6 teslas, moving at a speed of 100 m/s (Fig. 2.29). Calculate the voltage induced in each conductor.

Solution

According to Eq. 2-25, we find

$$E = Blv$$

$$= 0.6 \times 2 \times 100 = 120\text{ V}$$

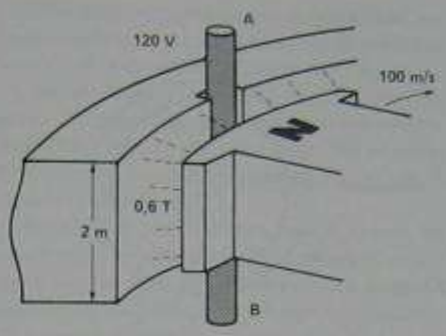


Figure 2.29
Voltage induced in a stationary conductor. See Example 2-9.

2.22 Lorentz force on a conductor

When a current-carrying conductor is placed in a magnetic field, it is subjected to a force which we call *electromagnetic force*, or Lorentz force. This force is of fundamental importance because it constitutes the basis of operation of motors, of generators, and of many electrical instruments. The magnitude of the force depends upon the orientation of the conductor with respect to the direction of the field. The force is greatest when the conductor is perpendicular to the field (Fig. 2.30) and zero when it is parallel to it (Fig. 2.31). Between these two extremes, the force has intermediate values.

The maximum force acting on a straight conductor is given by

$$F = BlI \tag{2.26}$$

where

F = force acting on the conductor [N]
B = flux density of the field [T]
l = active length of the conductor [m]
I = current in the conductor [A]

Example 2-10

A conductor 3 m long carrying a current of 200 A is placed in a magnetic field whose density is 0.5 T. Calculate the force on the conductor if it is perpendicular to the lines of force (Fig. 2.30).

Solution

$$F = BlI$$
$$= 0.5 \times 3 \times 200 = 300 \text{ N}$$

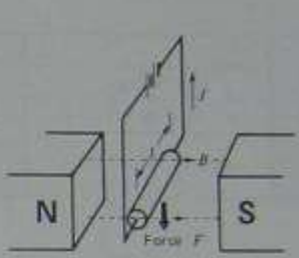


Figure 2.30
Force on a conductor.

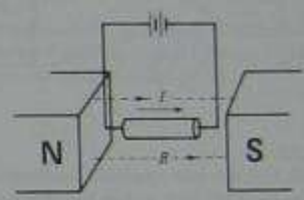


Figure 2.31
Force = 0.

2.23 Direction of the force acting on a straight conductor

Whenever a conductor carries a current, it is surrounded by a magnetic field. For a current flowing into the page of this book, the circular lines of force have the direction shown in Figure 2.32a. The same figure shows the magnetic field created between the N, S poles of a powerful permanent magnet.

The magnetic field does not, of course, have the shape shown in the figure because lines of force never cross each other. What, then, is the shape of the resulting field?

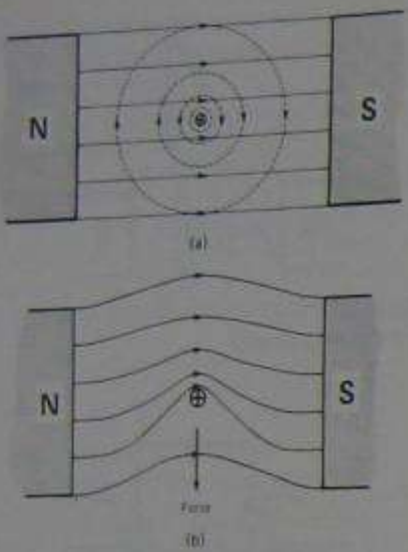


Figure 2.32
a. Magnetic field due to magnet and conductor.
b. Resulting magnetic field pushes the conductor downward.

To answer the question, we observe that the lines of force created respectively by the conductor and the permanent magnet act in the same direction above the conductor and in opposite directions below it. Consequently, the number of lines above the conductor must be greater than the number below. The resulting magnetic field therefore has the shape given in Figure 2.32b.

Recalling that lines of flux act like stretched elastic bands, it is easy to visualize that a force acts upon the conductor, tending to push it downward.

2.24 Residual flux density and coercive force

Consider the coil of Figure 2.33a, which surrounds a magnetic material formed in the shape of a ring. A current source, connected to the coil, produces a current whose value and direction can be changed at will. Starting from zero, we gradually increase I, so that H and B increase. This increase traces out

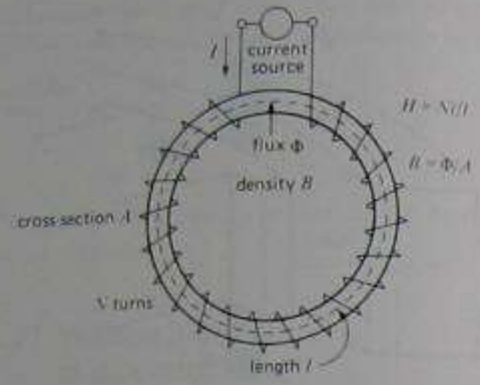


Figure 2.33a
Method of determining the *B-H* properties of a magnetic material.

curve **oa** in Figure 2.33b. The flux density reaches a value B_m for a magnetic field intensity H_m.

If the current is now gradually reduced to zero, the flux density B does not follow the original curve, but moves along a curve **ab** situated above **oa**. In effect, as we reduce the magnetic field intensity, the magnetic domains that were lined up under the influence of field H_m tend to retain their original orientation. This phenomenon is called hysteresis. Consequently, when H is reduced to zero, a substantial flux density remains. It is called residual flux density or *residual induction* (B_r).

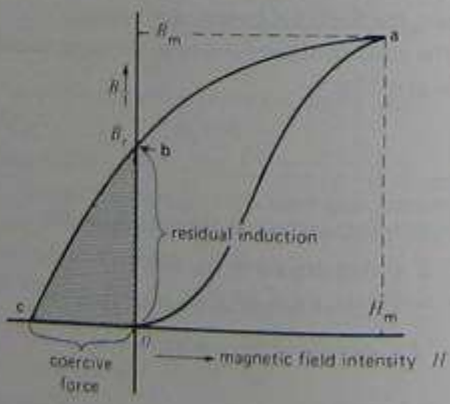


Figure 2.33b
Residual induction and coercive force.

If we wish to eliminate this residual flux, we have to reverse the current in the coil and gradually increase H in the opposite direction. As we do so, we move along curve **bc**. The magnetic domains gradually change their previous orientation until the flux density becomes zero at point **c**. The magnetic field intensity required to reduce the flux to zero is called *coercive force* (H_c).

In reducing the flux density from B_r to zero, we also have to furnish energy. This energy is used to overcome the frictional resistance of the magnetic domains as they oppose the change in orientation. The energy supplied is dissipated as heat in the material. A very sensitive thermometer would indicate a slight temperature rise as the ring is being demagnetized.

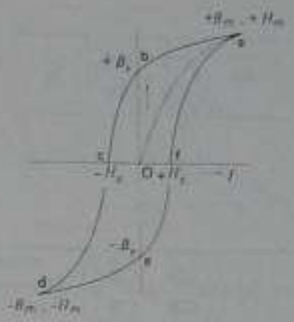


Figure 2.34
Hysteresis loop. If B is expressed in teslas and H in amperes per meter, the area of the loop is the energy dissipated per cycle, in joules per kilogram.

2.25 Hysteresis loop

Transformers and most electric motors operate on alternating current. In such devices the flux in the iron changes continuously both in value and direction. The magnetic domains are therefore oriented first in one direction, then the other, at a rate that depends upon the frequency. Thus, if the flux has a frequency of 60 Hz, the domains describe a complete cycle every 1/60 of a second, passing successively through peak flux densities $+B_m$ and $-B_m$ as the peak magnetic field intensity alternates between $+H_m$ and $-H_m$. If we plot the flux density B as a function of H, we obtain a closed curve called hysteresis loop (Fig. 2.34). The residual induction B_r and coercive force H_c have the same significance as before.

2.26 Hysteresis loss

In describing a hysteresis loop, the flux moves successively from $+B_m$, $+B_r$, 0, $-B_m$, $-B_r$, 0, and $+B_m$, corresponding respectively to points **a**, **b**, **c**, **d**, **e**, **f**, and **a**, of Figure 2.34. The magnetic material absorbs energy during each cycle and this energy is dissipated as heat. We can prove that the amount of heat released per cycle (expressed in J/m^3) is equal to the area (in T·A/m) of the hysteresis loop.

To reduce hysteresis losses, we select magnetic materials that have a narrow hysteresis loop, such as the grain-oriented silicon steel used in the cores of alternating-current transformers.

2.27 Hysteresis losses caused by rotation

Hysteresis losses are also produced when a piece of iron rotates in a constant magnetic field. Consider, for example, an armature AB, made of iron, that revolves in a field produced by permanent magnets N, S (Fig. 2.35). The magnetic domains in the armature tend to line up with the magnetic field, irrespective of the position of the armature. Consequently, as the armature rotates, the N poles of the domains point first toward A and then toward B. A complete reversal occurs therefore every half-revolution, as can be seen in Fig. 2.35a and 2.35b. Consequently, the magnetic domains in the armature reverse periodically, even though the magnetic field is constant. Hysteresis losses are produced just as they are in an ac magnetic field.

Learning Outcome 1.1

State the necessary conditions to produce a rotating magnetic field from stationary coils energised with an AC Supply

10.11 Rotating Field due to a Three-phase Winding

Fig. 10.17 shows a stator winding with three diametral coils aa', bb' and cc', each having N_s turns. The dots and crosses indicate the direction of conventionally positive current in each coil as explained

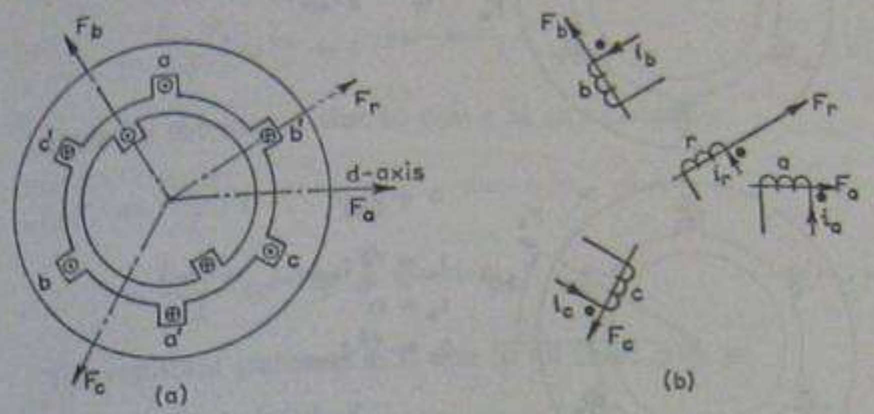


Fig. 10.17 M.M.F. DUE TO A 3-PHASE WINDING

in Section 10.3. The axes of the coil m.m.f.s are therefore mutually displaced by $2\pi/3$ radians, as shown in Fig. 10.17.

Suppose the three coils are supplied with balanced 3-phase currents, i_a, i_b and i_c, such that

$$i_a = I_{sm} \cos \omega t = \frac{I_{sm}}{2}(e^{j\omega t} + e^{-j\omega t}) \tag{10.44}$$

$$i_b = I_{sm} \cos (\omega t - 2\pi/3) = \frac{I_{sm}}{2}(e^{j(\omega t - 2\pi/3)} + e^{-j(\omega t - 2\pi/3)}) \tag{10.45}$$

$$i_c = I_{sm} \cos (\omega t + 2\pi/3) = \frac{I_{sm}}{2}(e^{j(\omega t + 2\pi/3)} + e^{-j(\omega t + 2\pi/3)}) \tag{10.46}$$

The m.m.f. of coil a is directed in the reference direction when i_a is positive. The instantaneous value of this m.m.f. is therefore

$$F_a' = \frac{I_{sm}N_s}{2}(e^{j\omega t} + e^{-j\omega t})e^{j0} \qquad (10.47)^*$$

This expression has been multiplied by e^{j0} $(= 1)$ to indicate that it acts in the space reference direction.

* To avoid confusion with f for frequency, instantaneous m.m.f. will be represented by F'.

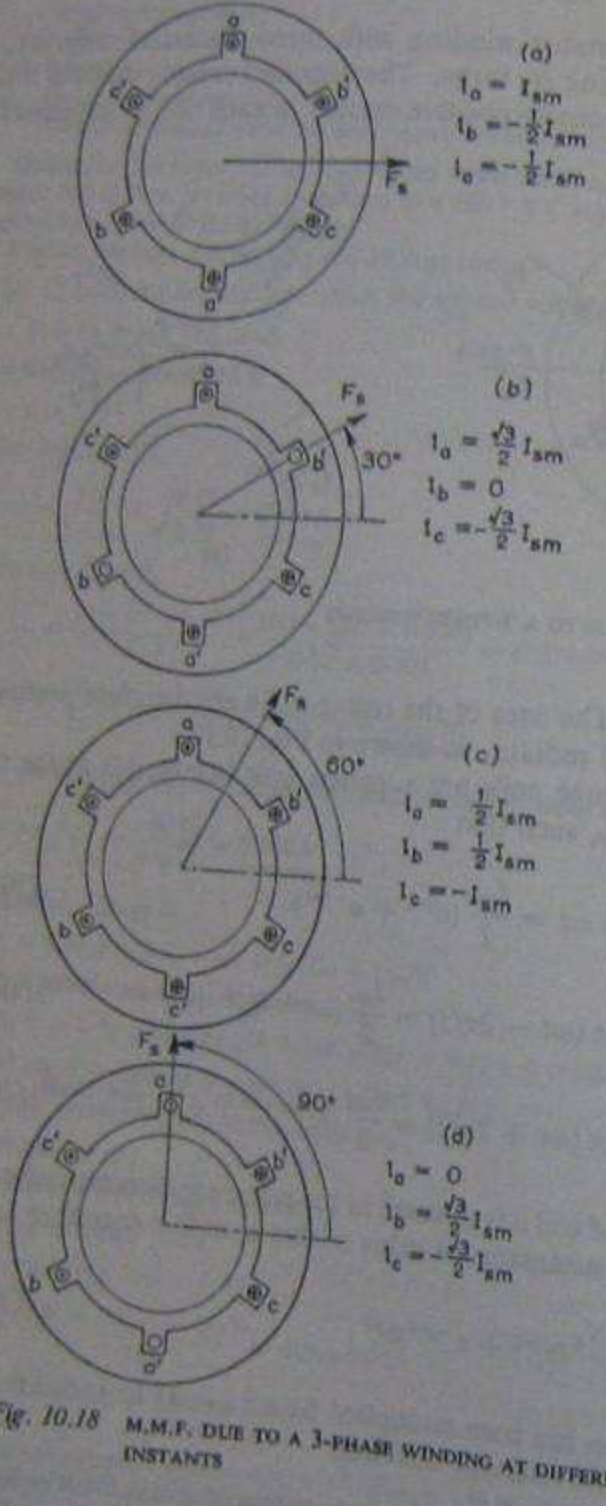


Fig. 10.18 M.M.F. DUE TO A 3-PHASE WINDING AT DIFFERENT INSTANTS

The m.m.f. of coil b is directed along an axis $+2\pi/3$ radians from the reference direction when i_b is positive. The instantaneous value of this m.m.f. is therefore

$$F_b' = \frac{I_{sm}N_s}{2}(e^{j(\omega t - 2\pi/3)} + e^{-j(\omega t - 2\pi/3)})e^{j2\pi/3}$$

$$= \frac{I_{sm}N_s}{2}(e^{j\omega t} + e^{-j(\omega t - 4\pi/3)}) \tag{10.48}$$

Similarly the m.m.f. due to coil c at any instant is

$$F_c' = \frac{I_{sm}N_s}{2}(e^{j(\omega t + 2\pi/3)} + e^{-j(\omega t + 2\pi/3)})e^{-j2\pi/3}$$

$$= \frac{I_{sm}N_s}{2}(e^{j\omega t} + e^{-j(\omega t + 4\pi/3)}) \tag{10.49}$$

The resultant stator m.m.f. due to all three coils is

$$F_s' = F_a' + F_b' + F_c'$$

$$= \frac{I_{sm}N_s}{2}[e^{j\omega t} + e^{-j\omega t} + e^{j\omega t} + e^{-j(\omega t - 4\pi/3)} + e^{j\omega t} + e^{-j(\omega t + 4\pi/3)}]$$

Since $e^{-j\omega t} + e^{-j(\omega t - 4\pi/3)} + e^{-j(\omega t + 4\pi/3)} = 0$,

$$F_s' = \tfrac{3}{2} I_{sm}N_s e^{j\omega t} \tag{10.50}$$

This equation shows that, when three coils are so positioned that their m.m.f. axes are mutually displaced by $2\pi/3$ radians and are then supplied with balanced 3-phase currents, an m.m.f. of constant magnitude results and the m.m.f. axis rotates at an angular velocity of ω radians per second.

For the coil configuration and phase sequence chosen the direction of rotation is in the $+\theta$ direction. It will be found that, if the phase sequence is reversed, the direction of rotation of the resultant m.m.f. axis is also reversed.

Fig. 10.18 shows the m.m.f. due to a 3-phase winding supplied with balanced 3-phase currents for a number of different instants. At (a) the current in phase a is positive maximum value and the currents in the two other phases are half the negative maximum value. The negative currents are indicated by showing the current in the cross direction in coil sides b and c, and in the dot direction in coil sides b' and c'. F_s is shown acting along the stator m.m.f. axis.

Figs. 10.18(b), (c) and (d) show successive instants in the 3-phase cycle corresponding to 30° rotations of the complexor diagram.

Learning Outcome 1.2

Calculate the synchronous speed of the rotating magnetic field given the supply frequency and number poles

13.2 Principle of operation

The operation of a 3-phase induction motor is based upon the application of Faraday's Law and the Lorentz force on a conductor (Sections 2.20, 2.21, and 2.22). The behavior can readily be understood by means of the following example.

Consider a series of conductors of length *l*, whose extremities are short-circuited by two bars A and B (Fig. 13.5a). A permanent magnet placed above this conducting ladder, moves rapidly to the right at a speed *v*, so that its magnetic field *B* sweeps across the conductors. The following sequence of events then takes place:

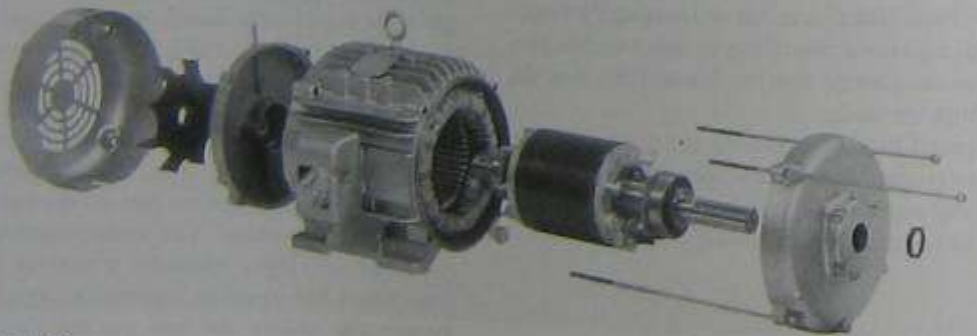


Figure 13.2
Exploded view of the cage motor of Fig. 13.1, showing the stator, rotor, end-bells, cooling fan, ball bearings, and terminal box. The fan blows air over the stator frame, which is ribbed to improve heat transfer. *(Courtesy of Baldor Electric Company)*

Figure 13.3a
Die-cast aluminum squirrel-cage rotor with integral cooling fan.
(Courtesy of Lab-Volt)

1. A voltage $E = Blv$ is induced in each conductor while it is being cut by the flux (Faraday's law).
2. The induced voltage immediately produces a current I, which flows down the conductor underneath the pole-face, through the end-bars, and back through the other conductors.
3. Because the current-carrying conductor lies in the magnetic field of the permanent magnet, it experiences a mechanical force (Lorentz force).
4. The force always acts in a direction to drag the conductor along with the magnetic field (Section 2.23).

If the conducting ladder is free to move, it will accelerate toward the right. However, as it picks up speed, the conductors will be cut less rapidly by the moving magnet, with the result that the induced voltage E and the current I will diminish. Consequently, the force acting on the conductors will also decrease. If the ladder were to move at the same speed as the magnetic field, the induced voltage E, the current I, and the force dragging the ladder along would all become zero.

In an induction motor the ladder is closed upon itself to form a squirrel-cage (Fig. 13.5b) and the moving magnet is replaced by a rotating field. The field is produced by the 3-phase currents that flow in the stator windings, as we will now explain.

13.3 The rotating field

Consider a simple stator having 6 salient poles, each of which carries a coil having 5 turns (Fig. 13.6). Coils that are diametrically opposite are connected in series by means of three jumpers that respectively connect terminals a-a, b-b, and c-c. This creates three identical sets of windings AN, BN, CN, that are mechanically spaced at 120° to each other. The

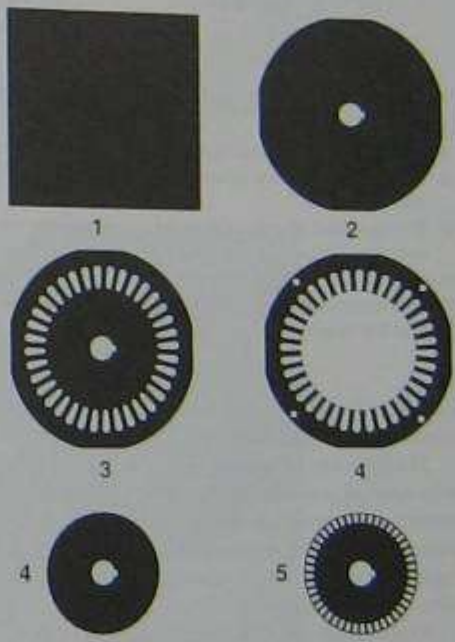


Figure 13.3b
Progressive steps in the manufacture of stator and rotor laminations. Sheet steel is sheared to size (1), blanked (2), punched (3), blanked (4), and punched (5).
(Courtesy of Lab-Volt)

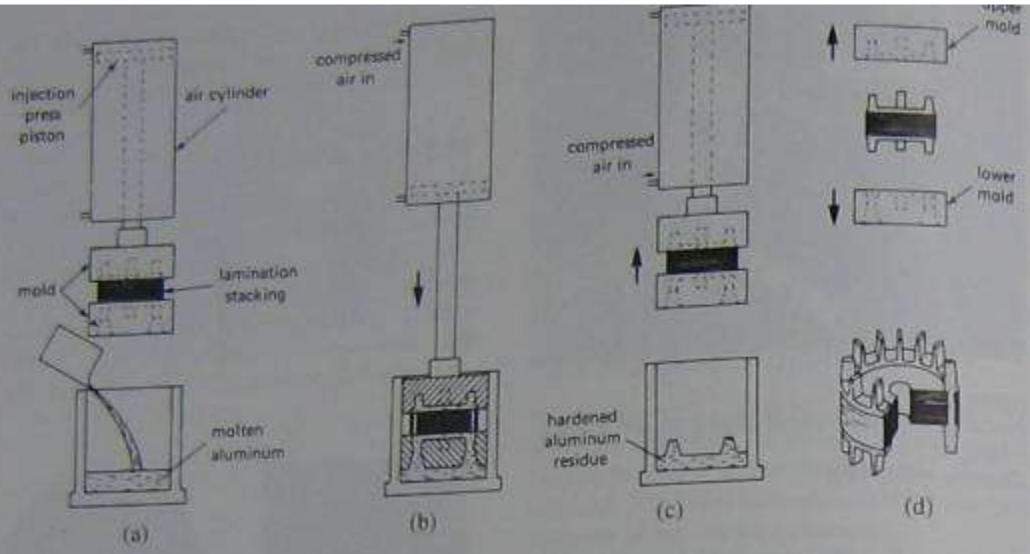


Figure 13.3c
Progressive steps in the injection molding of a squirrel-cage rotor.

a. Molten aluminum is poured into a cylindrical cavity. The laminated rotor stacking is firmly held between two molds.
b. Compressed air rams the mold assembly into the cavity. Molten aluminum is forced upward through the rotor bar holes and into the upper mold.
c. Compressed air withdraws the mold assembly, now completely filled with hot (but hardened) aluminum.
d. The upper and lower molds are pulled away, revealing the die-cast rotor. The cross section view shows that the upper and lower end-rings are joined by the rotor bars. *(Lab-Volt)*

two coils in each winding produce magnetomotive forces that act in the same direction.

The three sets of windings are connected in wye, thus forming a common neutral N. Owing to the perfectly symmetrical arrangement, the line-to-neutral impedances are identical. In other words, as regards terminals A, B, C, the windings constitute a balanced 3-phase system.

If we connect a 3-phase source to terminals A, B, C, alternating currents I_a, I_b, and I_c will flow in the windings. The currents will have the same value but will be displaced in time by an angle of 120°. These currents produce magnetomotive forces which, in turn, create a magnetic flux. It is this flux we are interested in.

In order to follow the sequence of events, we assume that positive currents (indicated by the arrows) always flow in the windings from line to neutral. Conversely, negative currents flow from neutral to line. Furthermore, to enable us to work with numbers, suppose that the peak current per phase is 10 A. Thus, when $I_a = +7$ A, the two coils of phase A will together produce an mmf of 7 A × 10 turns = 70 ampere-turns and a corresponding value of flux. Because the current is positive, the flux is directed vertically upward, according to the right-hand rule.

As time goes by, we can determine the instantaneous value and direction of the current in each winding and thereby establish the successive flux patterns. Thus, referring to Fig. 13.7 at instant 1, current I_a has a value of +10 A, whereas I_b and I_c both have a value of −5 A. The mmf of phase A is 10 A × 10 turns = 100 ampere-turns, while the mmf

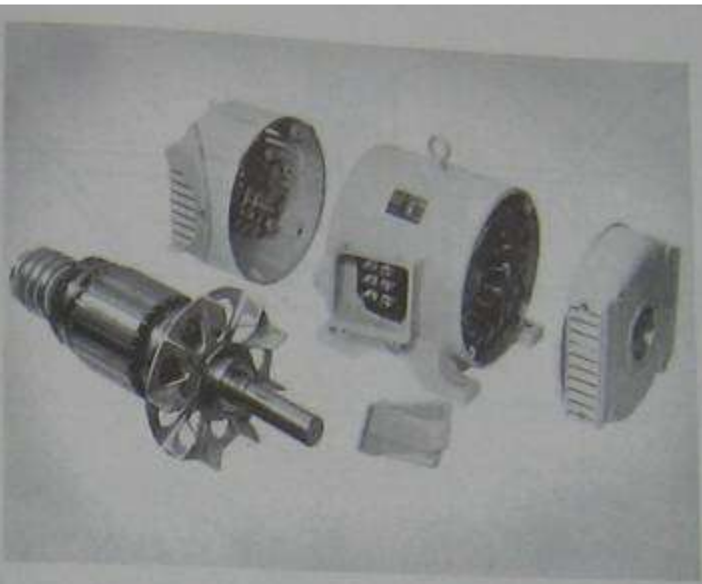


Figure 13.4a
Exploded view of a 5 hp, 1730 r/min wound-rotor induction motor.

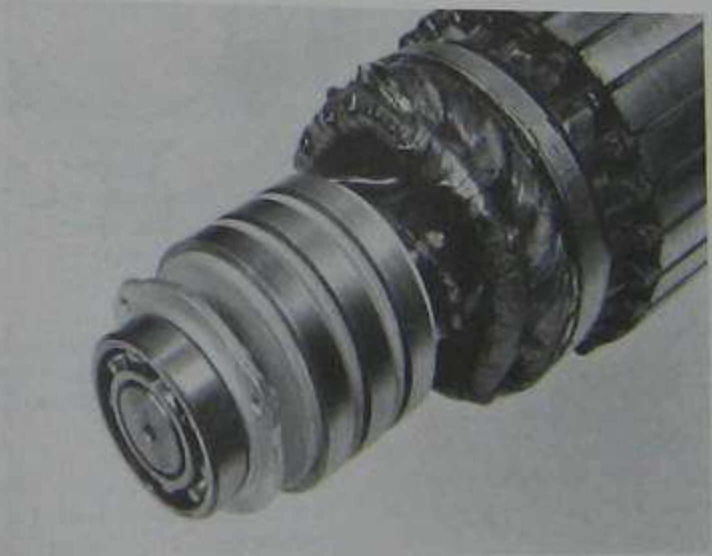


Figure 13.4b
Close-up of the slip-ring end of the rotor.
(Courtesy of Brook Crompton Parkinson Ltd)

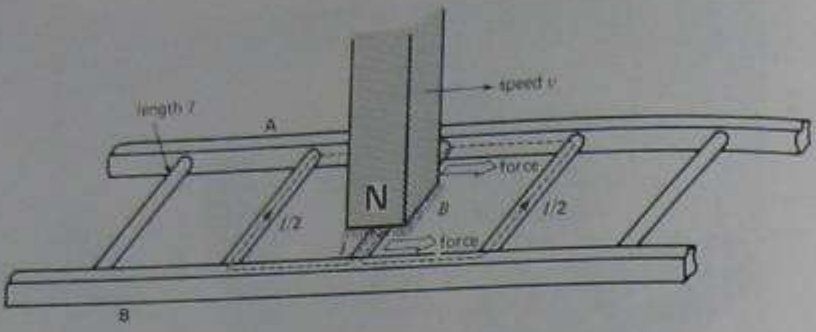


Figure 13.5a
Moving magnet cutting across a conducting ladder.

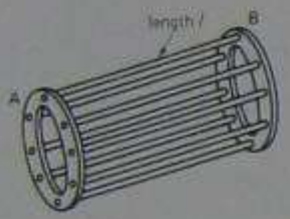


Figure 13.5b
Ladder bent upon itself to form a squirrel-cage.

of phases B and C are each 50 ampere-turns. The direction of the mmf depends upon the instantaneous current flows and, using the right-hand rule, we find that the direction of the resulting magnetic field is as shown in Fig. 13.8a. Note that as far as the rotor is concerned, the six salient poles together produce a magnetic field having essentially one broad north pole and one broad south pole. This means that the 6-pole stator actually produces a 2-pole field. The combined magnetic field points upward.

At instant 2, one-sixth cycle later, current I_c attains a peak of -10 A, while I_a and I_b both have a value of $+5$ A (Fig. 13.8b). We discover that the new field has the same shape as before, except that it has moved clockwise by an angle of 60°. In other words, the flux makes 1/6 of a turn between instants 1 and 2.

Proceeding in this way for each of the successive instants 3, 4, 5, 6, and 7, separated by intervals of 1/6

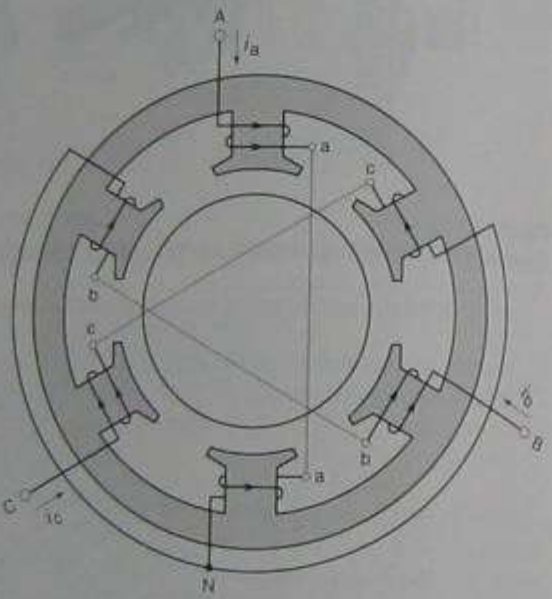


Figure 13.6
Elementary stator having terminals A, B, C connected to a 3-phase source (not shown). Currents flowing from line to neutral are considered to be positive.

cycle, we find that the magnetic field makes one complete turn during one cycle (see Figs. 13.8a to 13.8f).

The rotational speed of the field depends, therefore, upon the duration of one cycle, which in turn depends on the frequency of the source. If the frequency is 60 Hz, the resulting field makes one turn in 1/60 s, that is, 3600 revolutions per minute. On

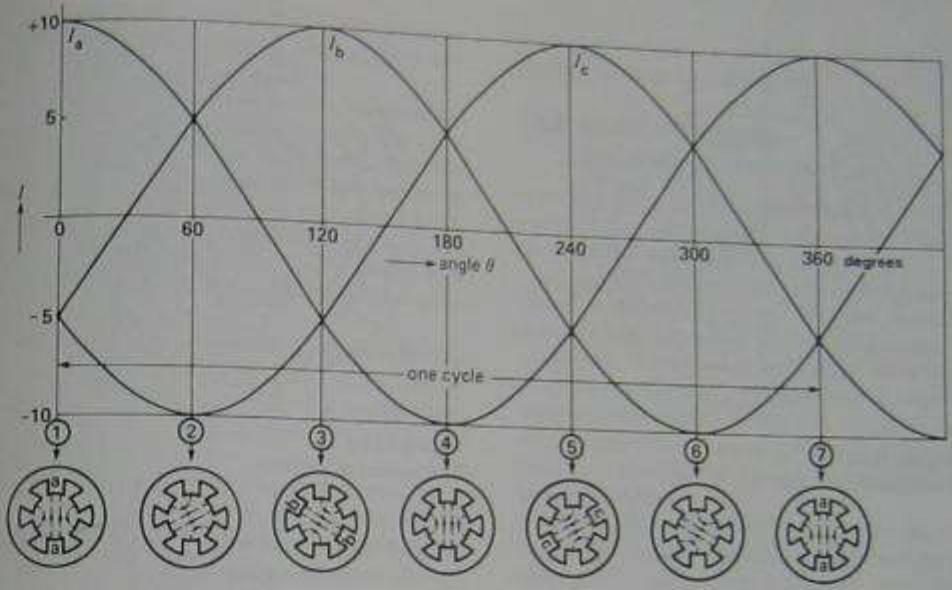


Figure 13.7
Instantaneous values of currents and position of the flux in Fig. 13.6.

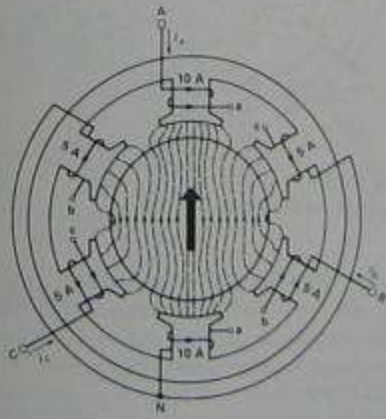


Figure 13.8a
Flux pattern at instant 1.

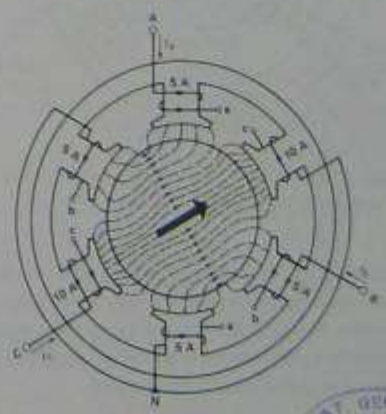


Figure 13.8b
Flux pattern at instant 2.

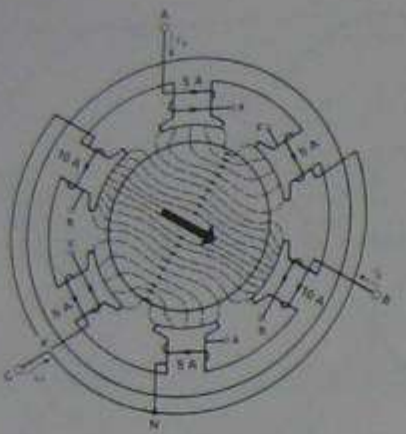


Figure 13.8c
Flux pattern at instant 3.

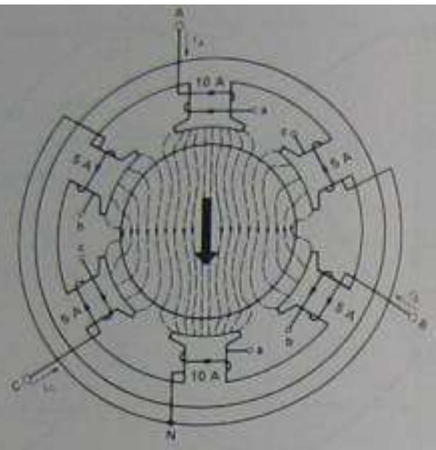


Figure 13.8d
Flux pattern at instant 4.

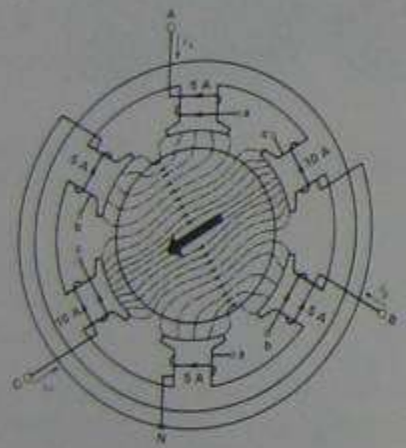


Figure 13.8e
Flux pattern at instant 5.

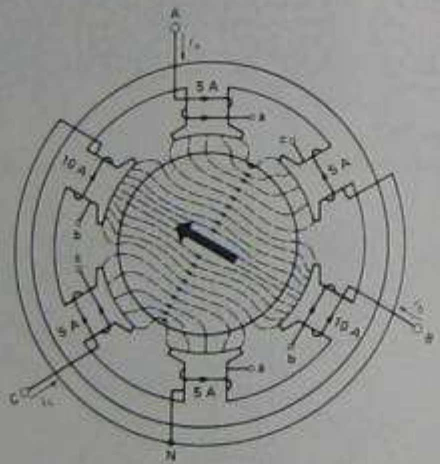


Figure 13.8f
Flux pattern at instant 6.

the other hand, if the frequency were 5 Hz, the field would make one turn in 1/5 s, giving a speed of only 300 r/min. Because the speed of the rotating field is necessarily synchronized with the frequency of the source, it is called *synchronous speed.*

13.4 Direction of rotation

The positive crests of the currents in Fig. 13.7 follow each other in the order A-B-C. This phase sequence produces a field that rotates clockwise. If we interchange any two of the lines connected to the stator, the new phase sequence will be A-C-B. By following the same line of reasoning developed in Section 13.3, we find that the field now revolves at synchronous speed in the opposite, or counterclockwise direction. Interchanging any two lines of a 3-phase motor will, therefore, reverse its direction of rotation.

Although early machines were built with salient poles, the stators of modern motors have internal diameters that are smooth. Thus, the salient-pole stator of Fig. 13.6 is now replaced by a smooth stator such as shown in Figs. 13.2 and 13.24a.

In Fig. 13.6, the two coils of phase A (Aa and An) are replaced by the two coils shown in Fig. 13.9a. They are lodged in two slots on the inner surface of the stator. Note that each coil covers 180° of the circumference whereas the coils in Fig. 13.6 cover only 60°. The 180° coil pitch is more efficient because it produces more flux per turn. A current I_a flowing from terminal A to the neutral N yields the flux distribution shown in the figure.

The coils of phases B and C are identical to those of phase A and, as can be seen in Fig. 13.9b, they are displaced at 120° to each other. The resulting magnetic field due to all three phases again consists of two poles.

In practice, instead of using a single coil per pole as shown in Fig. 13.9a, the coil is subdivided into two, three or more coils lodged in adjacent slots. The staggered coils are connected in series and constitute what is known as a *phase group*. Spreading the coil in this way over two or more slots tends to create a sinusoidal flux distribution per pole, which improves the performance of the motor and makes it less noisy. A phase group (or simply *group*) composed of 5 staggered coils connected in series to be placed in 5 successive slots is shown in Fig. 13.20.

13.5 Number of poles—synchronous speed

Soon after the invention of the induction motor, it was found that the speed of the revolving flux could be reduced by increasing the number of poles.

To construct a 4-pole stator, the coils are distributed as shown in Fig. 13.10a. The four identical groups of phase A now span only 90° of the stator circumference. The groups are connected in series and in such a way that adjacent groups produce magnetomotive forces acting in opposite directions. In other words, when a current I_a flows in the stator winding of phase A (Fig. 13.10a), it creates four alternate N-S poles.

The windings of the other two phases are identical but are displaced from each other (and from phase A) by a mechanical angle of 60°. When the wye-connected windings are connected to a 3-phase source, a revolving field having four poles is created (Fig. 13.10b). This field rotates at only half the speed of the 2-pole field shown in Fig. 13.9b. We will shortly explain why this is so.

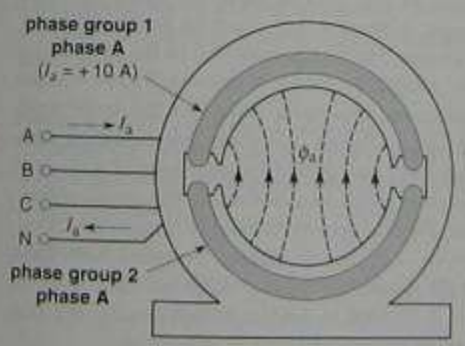


Figure 13.9a
Phase group 1 is composed of a single coil lodged in two slots. Phase group 2 is identical to Phase group 1. The two coils are connected in series. In practice, a phase group usually consists of two or more staggered coils.

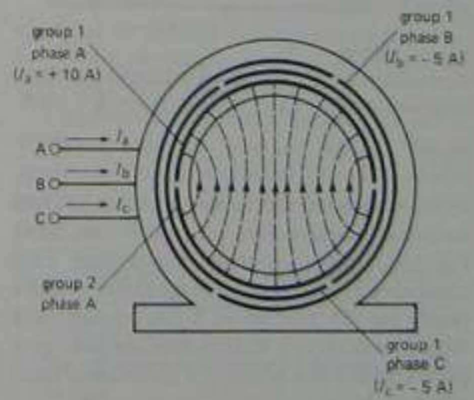


Figure 13.9b
Two-pole, full-pitch, lap-wound stator and resulting magnetic field when the current in phase A = +10 A and $I_b = I_c = -5$ A.

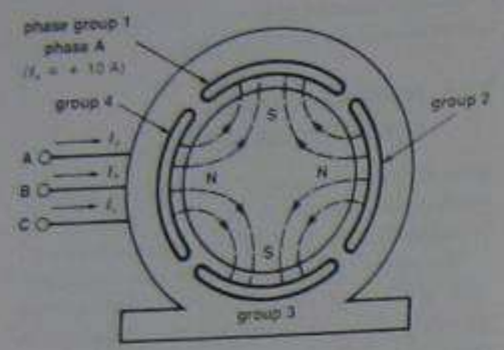


Figure 13.10a
The four phase groups of phase A produce a 4-pole magnetic field.

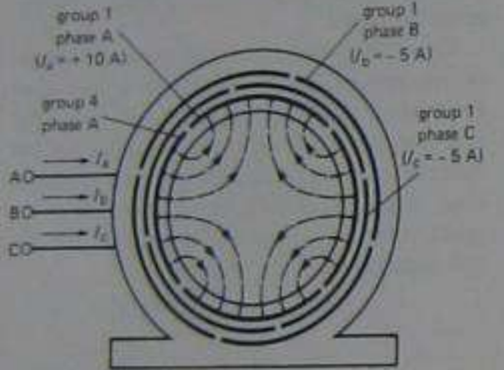


Figure 13.10b
Four-pole, full-pitch, lap-wound stator and resulting magnetic field when $I_a = +10$ A and $I_b = I_c = -5$ A.

We can increase the number of poles as much as we please provided there are enough slots. Thus, Fig. 13.11 shows a 3-phase, 8-pole stator. Each phase consists of 8 groups, and the groups of all the phases together produce an 8-pole rotating field. When connected to a 60 Hz source, the poles turn, like the spokes of a wheel, at a synchronous speed of 900 r/min.

How can we tell what the synchronous speed will be? Without going into all the details of current flow in the three phases, let us restrict our attention to phase A. In Fig. 13.11 each phase group covers a mechanical angle of 360/8 = 45°. Suppose the current in phase A is at its maximum positive value. The magnetic flux is then centered on phase A, and the N-S poles are located as shown in Fig. 13.12a. One-half cycle later, the current in phase A will reach its maximum negative value. The flux pattern will be the same as before, except that all the N poles will have become S poles and vice versa (Fig. 13.12b). In comparing the two figures, it is clear that the entire magnetic field has shifted by an angle of 45°—and this gives us the clue to finding the speed of rotation. The flux moves 45° and so it takes 8 half-cycles (= 4 cycles) to make a complete turn. On a 60 Hz system the time to make one turn is therefore 4 × 1/60 = 1/15 s. Consequently, the flux turns at the rate of 15 r/s or 900 r/min.

The speed of a rotating field depends therefore upon the frequency of the source and the number of poles on the stator. Using the same reasoning as above, we can prove that the synchronous speed is always given by the expression

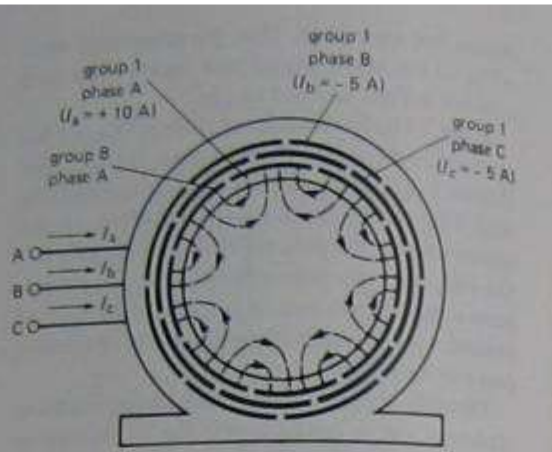


Figure 13.11
Eight-pole, full-pitch, lap-wound stator and resulting magnetic field when $I_a = +10$ A and $I_b = I_c = -5$ A.

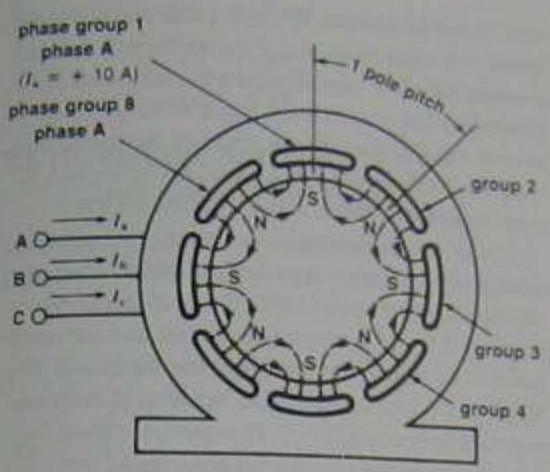


Figure 13.12a
Flux pattern when the current in phase A is at its maximum positive value.

$$n_s = \frac{120f}{p} \tag{13.1}$$

where

n_s = synchronous speed [r/min]
f = frequency of the source [Hz]
p = number of poles

This equation shows that the synchronous speed increases with frequency and decreases with the number of poles.

Example 13-1

Calculate the synchronous speed of a 3-phase induction motor having 20 poles when it is connected to a 50 Hz source.

Solution

$$n_s = 120f/p = 120 \times 50/20$$
$$= 300 \text{ r/min}$$

Learning Outcome 1.3

Determine the volts per turn, pitch and breadth factors given the desired value of stator flux density

THREE-PHASE WINDINGS AND FIELDS

In an a.c. machine the armature (or main) winding may be either on the stator (i.e. the stationary part of the machine) or on the rotor, the same form of winding being used in each case. The simplest form of 3-phase winding has concentrated coils each spanning one pole pitch, and with the starts of each spaced 120° (electrical) apart on the stator or rotor. These coils may be connected in star or delta as required.

In most machines the coils are not concentrated but are distributed in slots over the surface of the stator or rotor, and it is this type of winding which will now be considered. The same type of winding is common to both synchronous and asynchronous (induction) machines.

11.1 Flux Density Distributions

In all a.c. machines an attempt is made to secure a sinusoidal flux density distribution in the air-gap. This may be achieved approximately by the distribution of the winding in slots round the air-gap or by using salient poles with shaped pole shoes.

In Fig. 11.1(*a*) a section of a multipolar machine is shown. If the flux density in the air-gap is to be sinusoidally distributed, the flux density must be zero on the inter-polar axes such as OA, OC and OE, and maximum on the polar axes OB and OD. Since successive poles are of alternate north and south polarities, the maximum flux densities along OB and OD are oppositely directed. Thus a complete cycle of variation of the flux density takes place in a

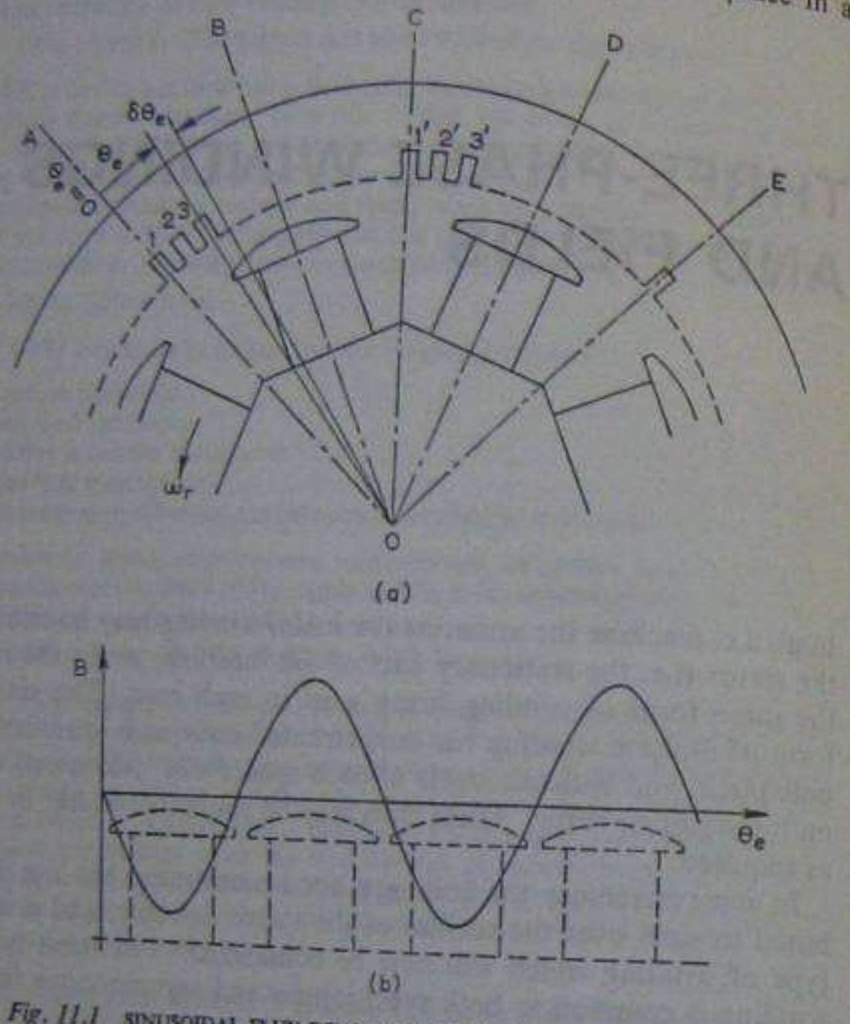


Fig. 11.1 SINUSOIDAL FLUX DENSITY DISTRIBUTION

double pole pitch from the axis OA to the axis OE. This is shown in Fig. 11.1(*b*).

Taking axis OA as the datum for angular measurements, the flux density at any point in the air-gap is

$$B = B_m \sin \theta_e \tag{11.1}$$

where θ_e is the angle from the origin measured in electrical radians or electrical degrees. Since one cycle of variation of the flux density occurs in a double pole pitch,

1 double pole pitch $\equiv 2\pi$ electrical radians or 360 electrical degrees

If the machine has $2p$ poles or p double pole pitches,

$$\theta_e = p\theta_m \tag{11.2}$$

where θ_m is the angular measure in mechanical radians or degrees.

11.2 Three-phase Single-layer Concentric Windings

The two sides of an armature coil must be placed in slots which are approximately a pole pitch (180 electrical degrees) apart so that the e.m.f.s in the coil sides are cumulative. In addition, in 3-phase machines the starts of each phase winding must be 120 electrical degrees apart.

In single-layer windings one coil side occupies the whole of a slot. As a result, difficulty is experienced in arranging the end connectors, or overhangs. In concentric and split-concentric windings differently shaped coils having different spans are necessary. To preserve e.m.f. balance in each of the phases, each phase must contain the same number of each shape of coil.

Fig. 11.2(*a*) represents a developed stator with 24 stator slots, and it is desired to place a 4-pole 3-phase concentric winding in them:

$$\text{Number of slots per pole} = \frac{24}{4} = 6$$

$$\text{Number of slots per pole and phase} = \frac{24}{4 \times 3} = 2$$

Fig. 11.2(*a*) shows the coil arrangement for the red phase as a thin full line. The start and finish (marked S and F respectively) of the phase winding are brought out, all the coils in the one phase being connected in series. For a phase sequence RYB, the yellow phase (shown dotted) must start 120 electrical degrees after the red phase. One pole pitch contains six slots and is equivalent to 180 electrical degrees. Hence a slot pitch is equivalent, in this case, to 30 electrical degrees.

The red phase starts in slot 1 and therefore the yellow phase must start in slot 5. In the same way the blue phase is 240 electrical degrees out of space phase with the red phase. The blue phase must therefore start in slot 9.

In Fig. 11.2 the finishes of the three phases have been commoned, making a star-connected winding. It would have been equally correct to common the three starts. The winding might also have been mesh-connected, in which case the finish of the red phase would have been connected to the start of the yellow phase, the finish of the yellow to the start of the blue, the finish of the blue to the start of the red, three connectors to the three junctions being brought out to terminals.

It will be observed that each phase has coils of each of the four different sizes used, thus maintaining balance between the phases.

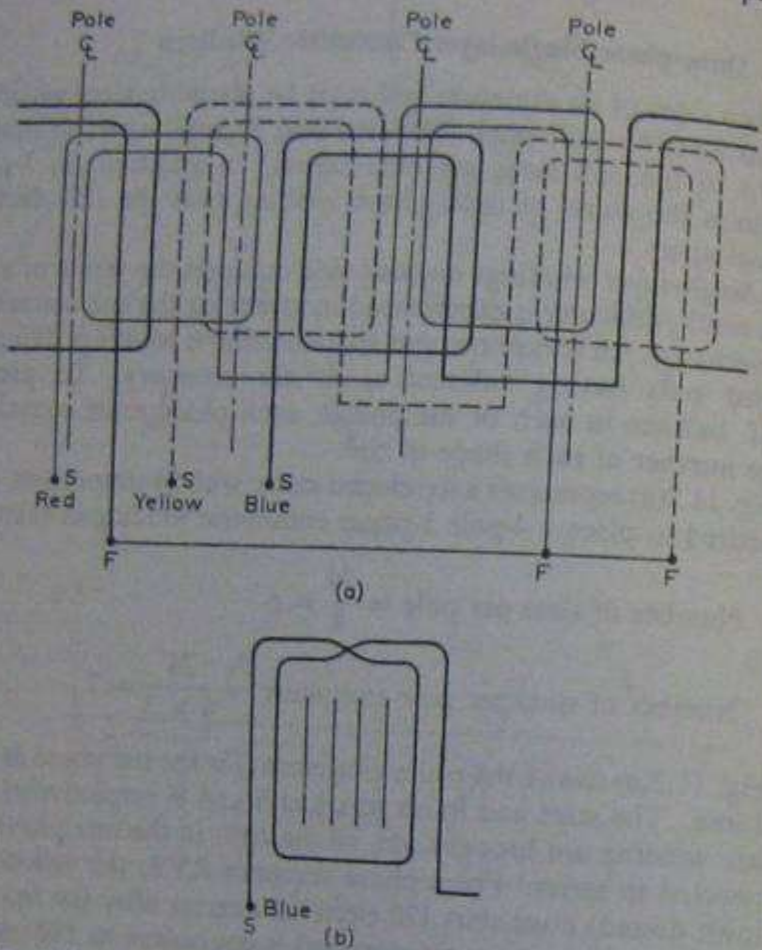


Fig. 11.2 FOUR-POLE 3-PHASE SINGLE-LAYER CONCENTRIC WINDING

It will also be seen that a coil group of any one phase consists of two coils per double pole pitch, one coil being greater than a pole pitch by one slot pitch and the other being less than a pole pitch by the same amount. If the end connexions of these two coils were crossed over as shown in Fig. 11.2(*b*) two full-pitch coils (i.e. having a span of exactly one-pole pitch) would be formed. Therefore each such coil group is the equivalent, electrically, of two full-pitch coils joined in series. All single-layer windings are effectively composed of full-pitch coils.

11.3 Three-phase Single-layer Mush Winding

Fig. 11.3 shows a 4-pole 3-phase single-layer mush winding. The distinctive feature of the mush winding is the utilization of constant-span coils. The overhangs are arranged in a similar manner to those of a conventional double-layer winding.

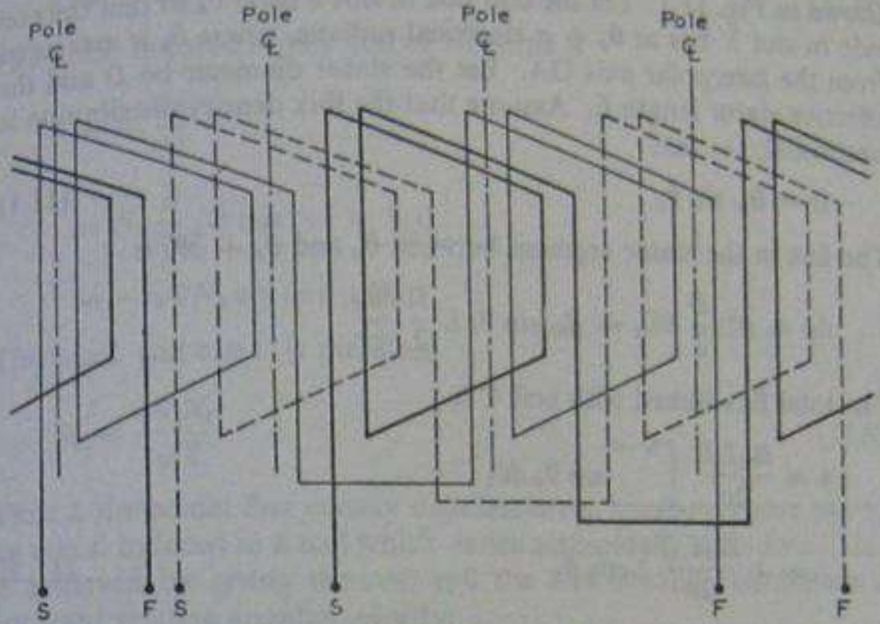


Fig. 11.3 FOUR-POLE 3-PHASE SINGLE-LAYER MUSH WINDING

11.4 Three-phase Double-layer Windings

The double-layer windings used in 3-phase machines are essentially similar to those used in d.c. machines except that no connexions to a commutator are required.

Since each phase must be balanced, all must contain equal numbers of coils and the starts of each phase must be displaced by 120 electrical degrees. If a number of groups of coils are to be connected in parallel, then similar parts in the winding at equal potentials must be available, a condition obtainable only in machines having a number of poles divisible by three when a wave winding is used.

On the other hand, tooth ripple, which arises where there are an integral number of slots per pole, resulting in the same relative positions of equivalent slots under each pole, may be avoided in double-layer windings by the use of winding pitches different from the pole pitch, thus giving a fractional number of slots per pole. A further advantage of the double-layer winding is the possibility of using constant-span coils. Only single-layer windings are considered in the rest of this chapter.

11.5 E.M.F. Induced in a Full-pitch Coil

Consider a full-pitch coil C with coil sides lying in slots 3 and 3′ as shown in Fig. 11.1. Let the coil side in slot 3 lie at θ_e so that the coil side in slot 3′ lies at $\theta_e + \pi$ electrical radians, where θ_e is measured from the interpolar axis OA. Let the stator diameter be D and the effective stator length L. Assume that the flux density distribution is sinusoidal, i.e. that

$$B = B_m \sin \theta_e \tag{11.1}$$

The flux in the stator segment between θ_e and $\theta_e + \delta\theta_e$ is

$$\delta\phi = BL\frac{D}{2}\,\delta\theta_m = B_m \sin \theta_e L\frac{D}{2}\frac{\delta\theta_e}{p}$$

The total flux linked with coil C is

$$\begin{aligned}\phi &= \frac{B_mLD}{2p}\int_{\theta_e}^{\theta_e+\pi} \sin \theta_e\, d\theta_e \\ &= +\frac{B_mLD}{2p}\,2\cos\theta_e \end{aligned}\tag{11.3}$$

If a coil lies with its sides on the interpolar axes, as, for example, the coil lying in slots 1 and 1′ of Fig. 11.1, then the coil links the total flux per pole, Φ:

$$\begin{aligned}\Phi &= \frac{B_mLD}{2p}\int_{0}^{\pi} \sin \theta_e\, d\theta_e \\ &= +\frac{B_mLD}{2p}\,2 \end{aligned}\tag{11.4}$$

The flux linked with coil C is therefore, by substitution in eqn. (11.3),

$$\phi = \Phi \cos \theta_e \tag{11.5}$$

Suppose the pole system rotates in the direction shown at a uniform angular velocity

$$\omega_r = 2\pi n_0 \quad \text{radians/second} \tag{11.6}$$

where n_0 is the rotor speed in revolutions per second. The position of any coil such as C at any instant, in electrical radians, is

$$\theta_e = \omega t + \theta_0$$

where θ_0 is the position of the coil at $t = 0$, and

$$\omega = p\omega_r = 2\pi n_0 p \quad \text{electrical radians/second} \tag{11.7}$$

Substituting for θ_e in eqn. (11.5), the flux linking any coil such as C at any time t is

$$\phi = \Phi \cos (\omega t + \theta_0) \tag{11.8}$$

The e.m.f. induced in any coil of N_c turns is

$$e = N_c \frac{d\phi}{dt}$$

$$= N_c \frac{d}{dt} \{\Phi \cos (\omega t + \theta_0)\}$$

$$= -\omega\Phi N_c \sin (\omega t + \theta_0)$$

The r.m.s. coil e.m.f. is therefore

$$E_c = \frac{\omega\Phi N_c}{\sqrt{2}} \tag{11.9}$$

Thus a sinusoidal flux density distribution in space may give rise to an e.m.f. induced in a coil which varies sinusoidally with time. This is achieved by giving the coil and the flux density distribution a constant relative angular velocity.

The frequency of the induced e.m.f. is

$$f = \frac{\omega}{2\pi} = \frac{2\pi n_0 p}{2\pi} = n_0 p \tag{11.10}$$

n_0 is called the *synchronous speed.* In this equation it is measured in revolutions per second.

11.6 Distribution (or Breadth) Factor and E.M.F. Equation

Suppose that under each pole pair each phase of the winding has g coils connected in series, each coil side being in a separate slot. The e.m.f. per phase and pole pair is the complexor sum of the coil voltages. These will not be in time phase with one another since successive coils are displaced round the armature, and hence will not be linked by the same value of flux at the same instant. $\boldsymbol{E}_1$, $\boldsymbol{E}_2$, $\boldsymbol{E}_3 \ldots \boldsymbol{E}_g$ (as shown in Fig. 11.4(*a*)) represent the r.m.s. values of the e.m.f.s in successive coils. The phase displacement between successive e.m.f.s is ψ, which depends on the electrical angular displacement between successive slots on the armature.

Suppose the machine has a total of S slots and $2p$ poles. Then

$$\text{Number of slots per pole} = \frac{S}{2p}$$

The slot pitch (electrical angle between slot centre lines) is

$$\psi = \frac{180°_e}{S/2p} \quad \text{(since 1 pole pitch} = 180°_e) \tag{11.11}$$

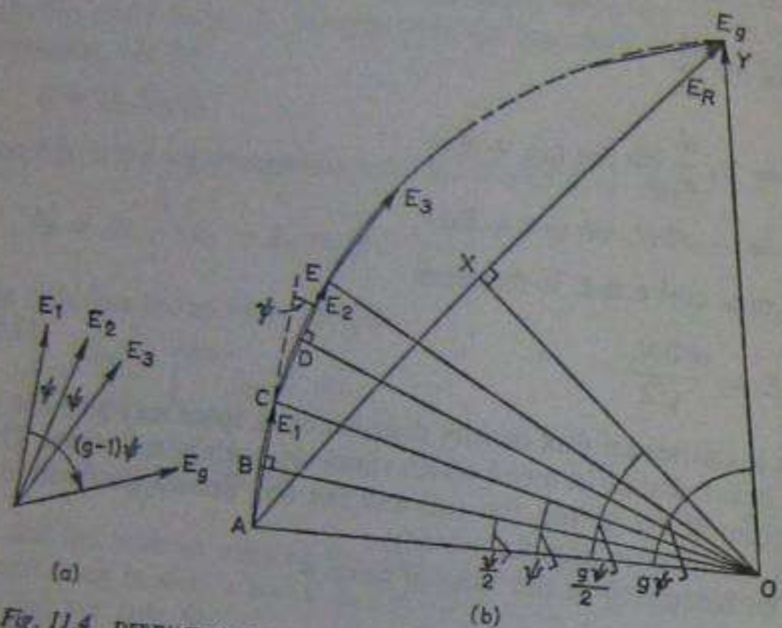


Fig. 11.4 DERIVATION OF DISTRIBUTION FACTOR
(a) Complexor diagram of slot e.m.f.s
(b) Resultant of slot e.m.f.s

The e.m.f. complexors $\boldsymbol{E}_1, \boldsymbol{E}_2, \boldsymbol{E}_3 \ldots \boldsymbol{E}_g$ are placed end to end in order in Fig. 11.4(*b*). The resultant complexor $\boldsymbol{E}_R$, represents the complexor sum of the e.m.f.s of the g coils connected in series.

Since the complexors $\boldsymbol{E}_1, \boldsymbol{E}_2, \boldsymbol{E}_3 \ldots \boldsymbol{E}_g$ are all of the same length and are displaced from one another by the same angle, they must be successive chords of the circle whose centre is O in Fig. 11.4(*b*). The complexor sum AY may be found as follows.

Join OA, OC, OE, etc., draw the perpendicular bisectors of each chord (i.e. OB, OD, etc.) and also the perpendicular bisector OX of the chord AY.

In the triangle AOX,

$$AX = AO \sin AOX = AO \sin g\frac{\psi}{2}$$

Therefore

$$AY = 2AO \sin g \frac{\psi}{2}$$

In the triangle AOB,

$$AB = AO \sin AOB = AO \sin \frac{\psi}{2}$$

$$AC = 2AB = 2AO \sin \frac{\psi}{2}$$

Therefore

$$\frac{AY}{AC} = \frac{E_R}{E_1} = \frac{\sin g \frac{\psi}{2}}{\sin \frac{\psi}{2}}$$

Thus the *distribution factor* is

$$K_d = \frac{\text{Complexor sum of coil e.m.f.s}}{\text{Arithmetic sum of coil e.m.f.s}}$$

$$= \frac{E_R}{gE_1} = \frac{\sin g \frac{\psi}{2}}{g \sin \frac{\psi}{2}} \tag{11.12}$$

The product $g\psi$ represents the electrical angle over which the conductors of one phase are spread under any one pole and is referred to as the *phase spread*. In a 3-phase single-layer winding each phase has two phase spreads under each pole pair. Therefore, for a single-layer 3-phase winding,

$$g\psi = \frac{360}{2 \times 3} = 60°_e \quad \text{or} \quad \pi/3 \text{ electrical radians}$$

Clearly the highest value which the distribution factor K_d can have is unity, corresponding to a situation where there is one coil per pole pair and phase. A lower limit for the value of K_d also exists. Thus, if the number of separate slots g in the phase spread $g\psi$ is considered to increase without limit, then

$$\psi \to 0 \quad \text{and} \quad \sin \frac{\psi}{2} \to \frac{\psi}{2}$$

A 3-phase winding with a phase spread of $60°_e$ is said to be *narrow spread*.

For a narrow-spread 3-phase winding ($g\psi = \pi/3$),

$$\lim_{\psi \to 0} K_d = \frac{\sin \frac{g\psi}{2}}{g \frac{\psi}{2}} = \frac{\sin \pi/6}{\pi/6} = \frac{3}{\pi} \tag{11.13}$$

A winding having this limiting condition is called a *uniform winding*, and in such winding the phase spreads may be thought of as current sheets with the effect of the slotting eliminated.

The lower limit of K_d for a 3-phase narrow-spread winding ($3/\pi = 0{\cdot}955$), corresponding to a very large number of slots per pole and phase, shows that the distribution of the winding will have little effect on the magnitude of the fundamental e.m.f. per phase.

Ideally the flux density distribution linking the winding should be sinusoidal. In practice this ideal is not usually achieved; the air-gap flux density distribution is then of the form

$$B = B_{m1} \sin \theta_e + B_{m3} \sin (3\theta_e + \epsilon_3) + \ldots B_{mn} \sin (n\theta_e + \epsilon_n) \tag{11.14}$$

In this expression the first term on the right-hand side is called the *fundamental space distribution*. The other terms are referred to as *space harmonics*. The *n*th space harmonic goes through n cycles of variation for one cycle of variation of the fundamental. Only odd space harmonics are present since the flux density distribution repeats itself under each pole and is therefore symmetrical.

Just as the fundamental flux density gives rise to a fundamental e.m.f. induced in a coil, so the *n*th space harmonic in the flux density distribution will give rise to an *n*th time harmonic in the coil e.m.f. The distribution factor for the *n*th harmonic is

$$K_{dn} = \frac{\sin \frac{gn\psi}{2}}{g \sin \frac{n\psi}{2}} \tag{11.15}$$

Although the distribution of the winding has little effect on the magnitude of the fundamental, it may cause considerable reduction in the magnitude of harmonic e.m.f.s compared with those occurring in a winding for which $g = 1$, i.e. one coil per pole pair and phase.

11.7 Coil-span Factor

The e.m.f. equation of Section 11.5 has been deduced on the assumption of full-pitch coils, i.e. coils whose sides are separated by one pole pitch. As has been pointed out, the coils in double-layer windings are often made either slightly more or slightly less than a pole pitch. Fig. 11.5 illustrates coils with various pitches.

If the coil has a pitch of exactly one pole pitch, it will at some instant link the entire flux of a rotor pole. If the coil pitch is less than one pole pitch, it will never link the entire flux of a rotor pole and the maximum coil e.m.f. will be reduced. If the coil pitch is greater than one pole pitch, the coil must always be linking flux

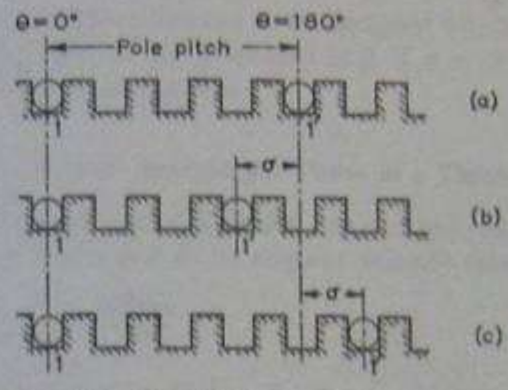


Fig. 11.5 COIL SPANS
(*a*) Full pitch
(*b*) Short pitch
(*c*) Over-full pitch

from at least two adjacent rotor poles so that the net flux linked will be less than the flux of one pole and the maximum coil e.m.f. will again be reduced.

The factor by which the e.m.f. per coil is reduced is called the *coil span factor*, K_s:

$$K_s = \frac{\text{E.M.F. in the short or long coil}}{\text{E.M.F. in a full-pitched coil}} \tag{11.16}$$

The magnitude of the coil span factor may most readily be obtained by considering the e.m.f. induced in each coil side, namely

$e = Blv$ volts

where B = air-gap flux density, l = active conductor length and v = conductor velocity at right angles to the direction of B.

This e.m.f. will have the same waveform as the flux density in the air-gap, since l and v are constant, and hence if the flux density is sinusoidally distributed the e.m.f. in each conductor will be sinusoidal so that the resultant coil e.m.f. will also be sinusoidal. If the pitch is short or long by an electrical angle σ, then, assuming a sinusoidal flux density distribution, the e.m.f.s in each side of the coil will differ in phase by σ but will have the same r.m.s. value. The resultant coil e.m.f. will be the complexor sum of the e.m.f.s in each coil side, as shown in Fig. 11.6.

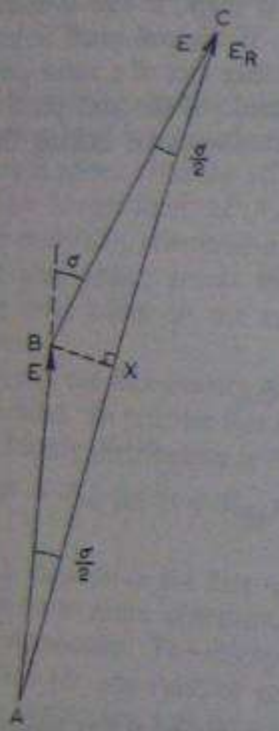


Fig. 11.6 DERIVATION OF COIL SPAN FACTOR

$$\text{Resultant e.m.f.} = AC = 2AB\cos\frac{\sigma}{2}$$

$$\text{E.M.F. for a full-pitch coil} = 2AB$$

Therefore

$$K_s = \frac{2AB\cos\frac{\sigma}{2}}{2AB} = \cos\frac{\sigma}{2} \tag{11.17}$$

If the flux density distribution contains space harmonics, the coil span factor for the nth harmonic e.m.f. is

$$K_{sn} = \cos\frac{n\sigma}{2} \tag{11.18}$$

All single-layer windings are effectively made up of full-pitch coils, but double-layer windings usually have short-pitched or short-chorded coils. The nth harmonic coil e.m.f. is reduced to zero if the *chording angle*, σ, is such that

$$\cos \frac{n\sigma}{2} = 0$$

or

$$\frac{n\sigma}{2} = 90°_e \tag{11.19}$$

This enables windings to be designed which will not permit specified harmonics to be generated (e.g. if $\sigma = 60°_e$ there can be no third-harmonic generation).

11.8 E.M.F. Induced per Phase of a Three-phase Winding

Following eqn. (11.9) the r.m.s. e.m.f. induced in a full-pitch coil of N_c turns due to its angular velocity relative to the pole system is

$$E_c = \frac{\omega \Phi N_c}{\sqrt{2}} \tag{11.9}$$

For a coil-span factor, K_s, due to chording,

$$E_c = K_s \frac{\omega \Phi N_c}{\sqrt{2}}$$

Further, if there are g coils in a phase group under a pole pair the resultant complexor sum is

$$E_g = K_d g E_c = K_d K_s g \frac{\omega \Phi N_c}{\sqrt{2}}$$

Assuming that the e.m.f.s of coil groups of the same phase under successive pole pairs are in phase and connected in series, the e.m.f. per phase is

$$E_p = pEg = pK_d K_s g \frac{\omega \Phi N_c}{\sqrt{2}}$$

or

$$E_p = K_d K_s \frac{\omega \Phi N_p}{\sqrt{2}} \tag{11.20}$$

where the number of turns per phase, N_p, is pgN_c.

This equation is sometimes written in the form

$$E_p = 4{\cdot}44\ K_d K_s f \Phi N_p \tag{11.21}$$

since $\omega = 2\pi f$ and $2\pi/\sqrt{2} = 4{\cdot}44$.

Learning Outcome 1.4

Determine the rotor impedance, current and power factor at a given value of slip

Example 13-6

A 3-phase, 8-pole squirrel-cage induction motor, connected to a 60 Hz line, possesses a synchronous speed of 900 r/min. The motor absorbs 40 kW, and the copper and iron losses in the stator amount to 5 kW and 1 kW, respectively. Calculate the torque developed by the motor.

Solution

The power transmitted across the air gap to the rotor is

$$\begin{aligned} P_r &= P_e - P_{js} - P_f \\ &= 40 - 5 - 1 = 34 \text{ kW} \end{aligned}$$

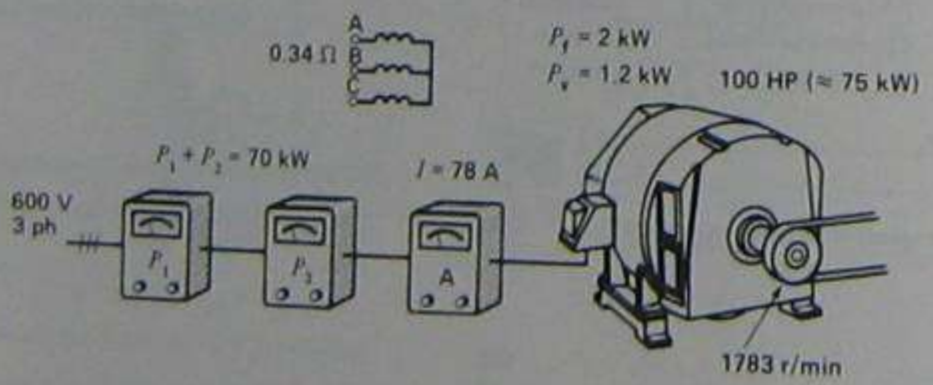
$$T_m = 9.55\, P_r / n_s \tag{13.9}$$

$$\begin{aligned} &= 9.55 \times 34\,000/900 \\ &= 361 \text{ N·m} \end{aligned}$$

Note that the solution to this problem (the torque) is independent of the speed of rotation. The motor could be at a standstill or running at full speed, but as long as the power P_r transmitted to the rotor is equal to 34 kW, the motor develops a torque of 361 N·m.

Example 13-7

A 3-phase induction motor having a nominal rating of 100 hp (~75 kW) and a synchronous speed of 1800 r/min is connected to a 600 V source (Fig. 13.16a). The two-wattmeter method shows a total power conwattmeter method shows a total pow



6a

le 13-7.

sumption of 70 kW, and an ammeter indicates a line current of 78 A. Precise measurements give a rotor speed of 1763 r/min. In addition, the following characteristics are known about the motor:

stator iron losses $P_f = 2$ kW
windage and friction losses $P_v = 1.2$ kW
resistance between two stator terminals $= 0.34\ \Omega$

Calculate

a. Power supplied to the rotor
b. Rotor I^2R losses
c. Mechanical power supplied to the load, in horsepower
d. Efficiency
e. Torque developed at 1763 r/min

Solution

a. Power supplied to the stator is

$$P_e = 70 \text{ kW}$$

Stator resistance per phase (assume a wye connection) is

$$R = 0.34/2 = 0.17\ \Omega$$

Stator I^2R losses are

$$P_{js} = 3I^2R = 3 \times (78)^2 \times 0.17 = 3.1 \text{ kW}$$

Iron losses $P_f = 2$ kW
Power supplied to the rotor:

$$P_r = P_e - P_{js} - P_f = (70 - 3.1 - 2) = 64.9 \text{ kW}$$

b. The slip is

$$s = (n_s - n)/n_s = (1800 - 1763)/1800 = 0.0205$$

Rotor I^2R losses:

$$P_{jr} = sP_r = 0.0205 \times 64.9 = 1.33 \text{ kW}$$

c. Mechanical power developed is

$$P_m = P_r - P_{jr} = 64.9 - 1.33 = 63.5 \text{ kW}$$

Mechanical power P_L to the load:

$$P_L = 63.5 - P_v = 63.5 - 1.2 = 62.3 \text{ kW} = 62.3 \times 1.34 \text{ (hp)} = 83.5 \text{ hp}$$

d. Efficiency of the motor is

$$\eta = P_L/P_e = 62.3/70 = 0.89 \text{ or } 89\%$$

e. Torque at 1763 r/min:

$$T = 9.55\ P_r/n_s = 9.55 \times 64\,900/1800 = 344 \text{ N·m}$$

The above calculations are summarized in Fig. 13.16b.

13.14 Torque versus speed curve

The torque developed by a motor depends upon its speed, but the relationship between the two cannot be expressed by a simple equation. Consequently, we prefer to show the relationship in the form of a

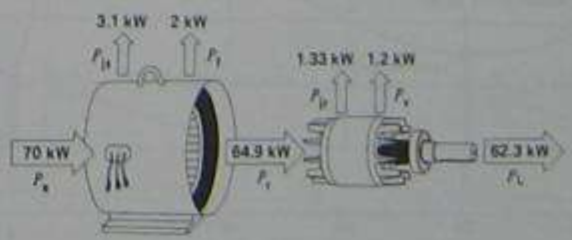


Figure 13.16b
Power flow in Example 13-7.

curve. Fig. 13.17 shows the torque-speed curve of a conventional 3-phase induction motor whose nominal full-load torque is T. The starting torque is 1.5 T and the maximum torque (called *breakdown torque*) is 2.5 T. Pull-up torque is the minimum torque developed by the motor while it is accelerating from rest to the breakdown torque.

At full-load the motor runs at a speed n. If the mechanical load increases slightly, the speed will drop until the motor torque is again equal to the load torque. As soon as the two torques are in balance, the motor will turn at a constant but slightly lower speed. However, if the load torque exceeds 2.5 T (the breakdown torque), the motor will quickly stop.

Small motors (15 hp and less) develop their breakdown torque at a speed n_d of about 80% of synchronous speed. Big motors (1500 hp and more) attain their breakdown torque at about 98% of synchronous speed.

13.15 Effect of rotor resistance

The rotor resistance of a squirrel-cage rotor is essentially constant from no-load to full-load, except that it increases with temperature. Thus, the resistance increases with increasing load because the temperature rises.

In designing a squirrel-cage motor, the rotor resistance can be set over a wide range by using copper, aluminum, or other metals in the rotor bars and end-rings. The torque-speed curve is greatly affected by such a change in resistance. The only characteristic that remains unchanged is the breakdown torque. The following example illustrates the changes that occur.

Figure 13.18a shows the torque-speed curve of a 10 kW (13.4 hp), 50 Hz, 380 V motor having a synchronous speed of 1000 r/min and a full-load torque of 100 N·m (~73.7 ft·lbf). The full-load current is 20 A and the locked-rotor current is 100 A. The rotor has an arbitrary resistance R.

Let us increase the rotor resistance by a factor of 2.5. This can be achieved by using a material of higher resistivity, such as bronze, for the rotor bars and end-rings. The new torque-speed curve is shown in Figure 13.18b. It can be seen that the starting torque doubles and the locked-rotor current decreases from 100 A to 90 A. The motor develops its breakdown torque at a speed N_d of 500 r/min, compared to the original breakdown speed of 800 r/min.

If we again double the rotor resistance so that it becomes 5 R, the locked-rotor torque attains a maximum value of 250 N·m for a corresponding current of 70 A (Fig. 13.18c).

A further increase in rotor resistance decreases both the locked-rotor torque and locked-rotor current. For example, if the rotor resistance is increased 25 times (25 R), the locked-rotor current drops to 20 A, but the motor develops the same

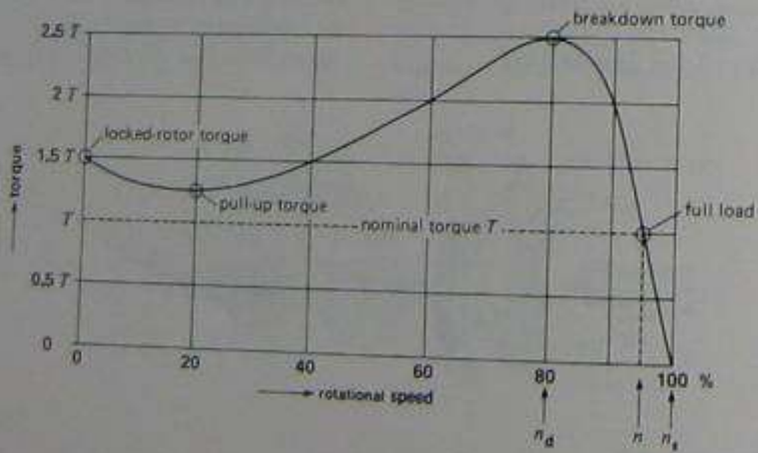


Figure 13.17
Typical torque-speed curve of a 3-phase squirrel-cage induction motor.

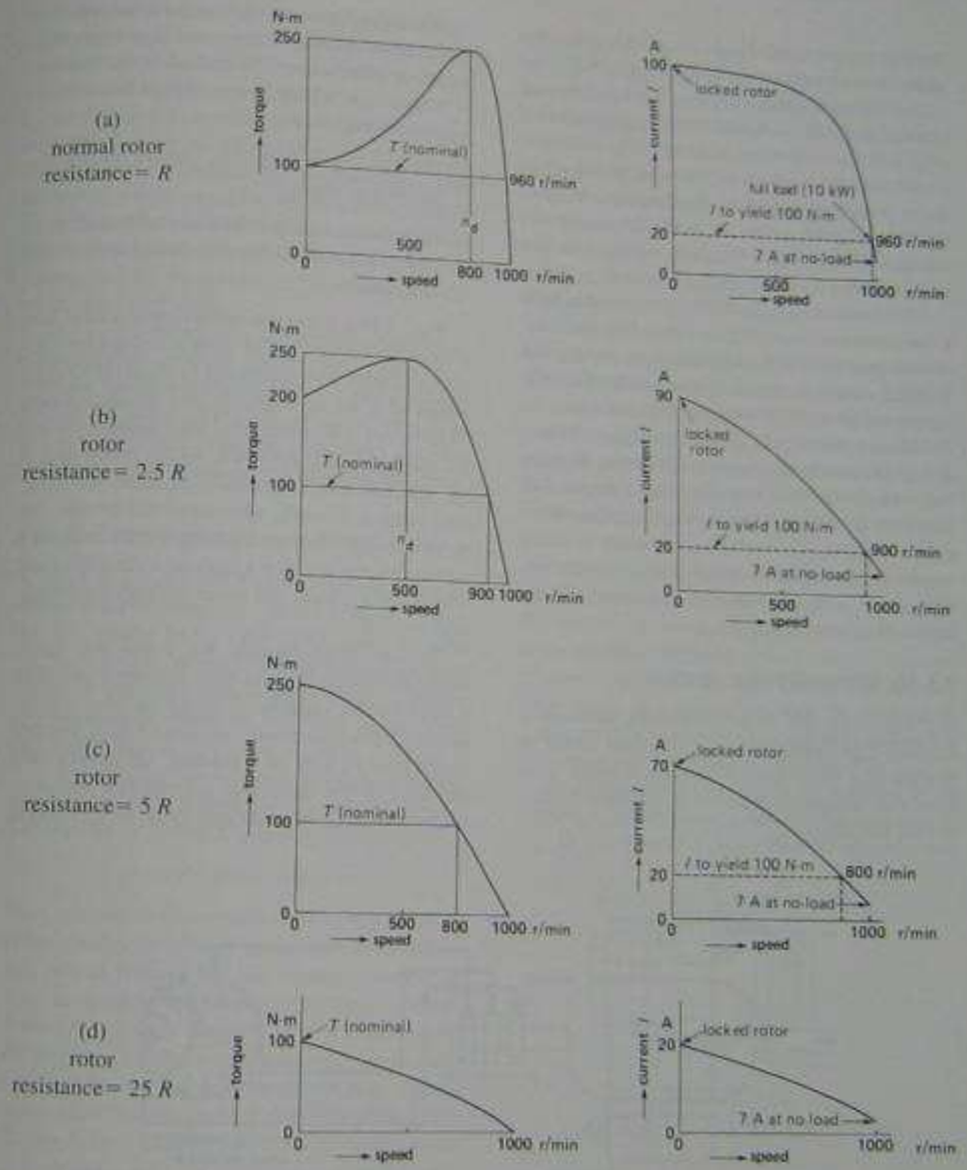


Figure 13.18
Rotor resistance affects the motor characteristics.

Learning Outcome 1.5

Calculate the rotor frequency given a value of supply frequency and slip

13.7 Acceleration of the rotor—slip

As soon as the rotor is released, it rapidly accelerates in the direction of the rotating field. As it picks up speed, the relative velocity of the field with respect to the rotor diminishes progressively. This causes both the value and the frequency of the induced voltage to decrease because the rotor bars are cut more slowly. The rotor current, very large at first, decreases rapidly as the motor picks up speed.

The speed will continue to increase, but it will never catch up with the revolving field. In effect, if the rotor *did* turn at the same speed as the field (synchronous speed), the flux would no longer cut the rotor bars and the induced voltage and current would fall to zero. Under these conditions the force acting on the rotor bars would also become zero and the friction and windage would immediately cause the rotor to slow down.

The rotor speed is always slightly less than synchronous speed so as to produce a current in the rotor bars sufficiently large to overcome the braking torque. At no-load the percent difference in speed between the rotor and field (called *slip*), is small: usually less than 0.1% of synchronous speed.

13.8 Motor under load

Suppose the motor is initially running at no-load. If we apply a mechanical load to the shaft, the motor will begin to slow down and the revolving field will cut the rotor bars at a higher and higher rate. The induced voltage and the resulting current in the bars will increase progressively, producing a greater and greater motor torque. The question is, for how long can this go on? Will the speed continue to drop until the motor comes to a halt?

No, the motor and the mechanical load will reach a state of equilibrium when the motor torque is exactly equal to the load torque. When this state is reached, the speed will cease to drop and the motor will turn at a constant rate. It is very important to understand that a motor only turns at constant speed when its torque is *exactly* equal to the torque exerted by the mechanical load. The moment this state of equilibrium is upset, the motor speed will start to change (Section 3.11).

Under normal loads, induction motors run very close to synchronous speed. Thus, at full-load, the slip for large motors (1000 kW and more) rarely exceeds 0.5% of synchronous speed, and for small machines (10 kW and less), it seldom exceeds 5%. That is why induction motors are considered to be constant speed machines. However, because they never actually turn at synchronous speed, they are sometimes called *asynchronous* machines.

13.9 Slip and slip speed

The slip s of an induction motor is the difference between the synchronous speed and the rotor speed, expressed as a percent (or per-unit) of synchronous speed. The per-unit slip is given by the equation

$$s = \frac{n_s - n}{n_s} \tag{13.2}$$

where

s = slip
n_s = synchronous speed [r/min]
n = rotor speed [r/min]

The slip is practically zero at no-load and is equal to 1 (or 100%) when the rotor is locked.

Example 13-2

A 0.5 hp, 6-pole induction motor is excited by a 3-phase, 60 Hz source. If the full-load speed is 1140 r/min, calculate the slip.

Solution

The synchronous speed of the motor is

$$n_s = 120f/p = 120 \times 60/6 \tag{13.1}$$
$$= 1200 \text{ r/min}$$

The difference between the synchronous speed of the revolving flux and rotor speed is the slip speed:

$$n_s - n = 1200 - 1140 = 60 \text{ r/min}$$

The slip is

$$s = (n_s - n)/n_s = 60/1200 \quad (13.2)$$
$$= 0.05 \text{ or } 5\%$$

13.10 Voltage and frequency induced in the rotor

The voltage and frequency induced in the rotor both depend upon the slip. They are given by the following equations:

$$f_2 = sf \quad (13.3)$$
$$E_2 = sE_{oc} \text{ (approx.)} \quad (13.4)$$

where

f_2 = frequency of the voltage and current in the rotor [Hz]

f = frequency of the source connected to the stator [Hz]

s = slip

E_2 = voltage induced in the rotor at slip s

E_{oc} = open-circuit voltage induced in the rotor when at rest [V]

In a cage motor, the open-circuit voltage E_{oc} is the voltage that *would* be induced in the rotor bars if the bars were disconnected from the end-rings. In the case of a wound-rotor motor the open-circuit voltage is $1/\sqrt{3}$ times the voltage between the open-circuit slip-rings.

It should be noted that Eq. 13.3 *always* holds true, but Eq. 13.4 is valid only if the revolving flux (expressed in webers) remains absolutely constant. However, between zero and full-load the actual value of E_2 is only slightly less than the value given by the equation.

Example 13-3

The 6-pole wound-rotor induction motor of Example 13-2 is excited by a 3-phase 60 Hz source. Calculate the frequency of the rotor current under the following conditions:

a. At standstill
b. Motor turning at 500 r/min in the same direction as the revolving field
c. Motor turning at 500 r/min in the opposite direction to the revolving field
d. Motor turning at 2000 r/min in the same direction as the revolving field

Solution

From Example 13-2, the synchronous speed of the motor is 1200 r/min.

a. At standstill the motor speed $n = 0$. Consequently, the slip is

$$s = (n_s - n)/n_s = (1200 - 0)/1200 = 1$$

The frequency of the induced voltage (and of the induced current) is

$$f_2 = sf = 1 \times 60 = 60 \text{ Hz}$$

b. When the motor turns in the same direction as the field, the motor speed n is positive. The slip is

$$s = (n_s - n)/n_s = (1200 - 500)/1200$$
$$= 700/1200 = 0.583$$

The frequency of the induced voltage (and of the rotor current) is

$$f_2 = sf = 0.583 \times 60 = 35 \text{ Hz}$$

c. When the motor turns in the opposite direction to the field, the motor speed is *negative*; thus, $n = -500$. The slip is

$$s = (n_s - n)/n_s$$
$$= [1200 - (-500)]/1200$$
$$= (1200 + 500)/1200 = 1700/1200$$
$$= 1.417$$

A slip greater than 1 implies that the motor is operating as a brake.

The frequency of the induced voltage and rotor current is

$$f_2 = sf = 1.417 \times 60 = 85 \text{ Hz}$$

d. The motor speed is positive because the rotor turns in the same direction as the field: $n = +2000$. The slip is

$$s = (n_s - n)/n_s$$
$$= (1200 - 2000)/1200$$
$$= -800/1200 = -0.667$$

A negative slip implies that the motor is actually operating as a generator.

The frequency of the induced voltage and rotor current is

$$f_2 = sf = -0.667 \times 60 = -40 \text{ Hz}$$

A negative frequency means that the phase sequence of the voltages induced in the rotor windings is reversed. Thus, if the phase sequence of the rotor voltages is A-B-C when the frequency is positive, the phase sequence is A-C-B when the frequency is negative. As far as a frequency meter is concerned, a negative frequency gives the same reading as a positive frequency. Consequently, we can say that the frequency is simply 40 Hz.

13.11 Characteristics of squirrel-cage induction motors

Table 13A lists the typical properties of squirrel-cage induction motors in the power range between 1 kW and 20 000 kW. Note that the current and torque are expressed in per-unit values. The base current is the full-load current and all other currents are compared to it. Similarly, the base torque is the full-load torque and all other torques are compared to it. Finally, the base speed is the synchronous speed of the motor. The following explanations will clarify the meaning of the values given in the table.

1. Motor at no-load. When the motor runs at no load, the stator current lies between 0.5 and 0.3 pu (of full-load current). The no-load current is similar to the exciting current in a transformer. Thus, it is composed of a magnetizing component that creates the revolving flux Φ_m and a small active component that supplies the windage and friction losses in the rotor plus the iron losses in the stator. The flux Φ_m links both the stator and the rotor; consequently it is similar to the mutual flux in a transformer (Fig. 13.13).

Considerable reactive power is needed to create the revolving field and, in order to keep it within acceptable limits, the air gap is made as short as mechanical tolerances will permit. The power factor at no-load is therefore low; it ranges from 0.2 (or 20%) for small machines to 0.05 for large machines. The efficiency is zero because the output power is zero.

2. Motor under load. When the motor is under load, the current in the rotor produces a mmf which tends to change the mutual flux Φ_m. This sets up an opposing current flow in the stator. The opposing mmfs of the rotor and stator are very similar to the opposing mmfs of the secondary and primary in a transformer. As a result, leakage fluxes Φ_{f1} and Φ_{f2} are created, in addition to the mutual flux Φ_m (Fig. 13.14). The total reactive power needed to produce these three fluxes is slightly greater than when the motor is operating at no-load. However, the active power (kW) absorbed by the motor increases in almost direct proportion to the mechanical load. It follows that the power factor

TABLE 13A TYPICAL CHARACTERISTICS OF SQUIRREL-CAGE INDUCTION MOTORS

Loading	Current (per-unit)		Torque (per-unit)		Slip (per-unit)		Efficiency		Power factor	
Motor size →	Small*	Big*	Small	Big	Small	Big	Small	Big	Small	Big
Full-load	1	1	1	1	0.03	0.004	0.7 to 0.9	0.96 to 0.98	0.8 to 0.85	0.87 to 0.9
No-load	0.5	0.3	0	0	~0	~0	0	0	0.2	0.05
Locked rotor	5 to 6	4 to 6	1.5 to 3	0.5 to 1	1	1	0	0	0.4	0.1

*Small means under 11 kW (15 hp); big means over 1120 kW (1500 hp) and up to 25 000 hp.

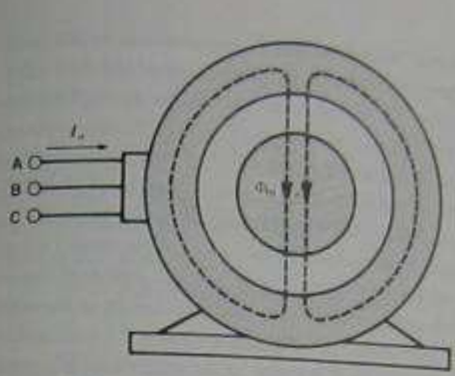


Figure 13.13
At no-load the flux in the motor is mainly the mutual flux Φ_m. To create this flux, considerable reactive power is needed.

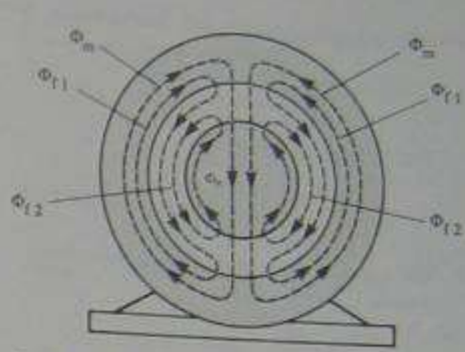


Figure 13.14
At full-load the mutual flux decreases, but stator and rotor leakage fluxes are created. The reactive power needed is slightly greater than in Fig. 13.13.

of the motor improves dramatically as the mechanical load increases. At full-load it ranges from 0.80 for small machines to 0.90 for large machines. The efficiency at full-load is particularly high; it can attain 98% for very large machines.

3. Locked-rotor characteristics. The locked-rotor current is 5 to 6 times the full-load current, making the I^2R losses 25 to 36 times higher than normal. The rotor must therefore never remain locked for more than a few seconds.

Although the mechanical power at standstill is zero, the motor develops a strong torque. The power factor is low because considerable reactive power is needed to produce the leakage flux in the rotor and stator windings. These leakage fluxes are much larger than in a transformer because the stator and the rotor windings are not as tightly coupled (see Section 10.2).

13.12 Estimating the currents in an induction motor

The full-load current of a 3-phase induction motor may be calculated by means of the following approximate equation:

$$I = 600\, P_h/E \qquad (13.5)$$

where

I = full-load current [A]
P_h = output power [horsepower]
E = rated line voltage (V)
600 = empirical constant

Recalling that the starting current is 5 to 6 pu and that the no-load current lies between 0.5 and 0.3 pu, we can readily estimate the value of these currents for any induction motor.

Example 13-4 ______

a. Calculate the approximate full-load current, locked-rotor current, and no-load current of a 3-phase induction motor having a rating of 500 hp, 2300 V.
b. Estimate the apparent power drawn under locked-rotor conditions.
c. State the nominal rating of this motor, expressed in kilowatts.

Solution

a. The full-load current is

$$I = 600\, P_h/E \qquad (13.5)$$
$$= 600 \times 500/2300$$
$$= 130 \text{ A (approx.)}$$

The no-load current is

$$I_o = 0.3I = 0.3 \times 130$$
$$= 39 \text{ A (approx.)}$$

The starting current is

$$I_{LR} = 6I = 6 \times 130$$
$$= 780 \text{ A (approx.)}$$

b. The apparent power under locked-rotor conditions is

$$S = \sqrt{3}\, EI$$
$$= \sqrt{3} \times 2300 \times 780 \qquad (8.9)$$
$$= 3100 \text{ kVA (approx.)}$$

c. When the power of a motor is expressed in kilowatts, it always relates to the mechanical output and *not* to the electrical input. The nominal rating of this motor expressed in SI units is, therefore,

$$P = 500/1.34$$
$$= 373 \text{ kW}$$ (see Power conversion chart in Appendix AX0)

13.13 Active power flow

Voltages, currents, and phasor diagrams enable us to understand the detailed behavior of an induction motor. However, it is easier to see how electrical energy is converted into mechanical energy by following the active power as it flows through the machine. Thus, referring to Fig. 13.15, active power P_e flows from the line into the 3-phase stator. Due to the stator copper losses, a portion P_{js} is dissipated as heat in the windings. Another portion P_f is dissipated as heat in the stator core, owing to the iron losses. The remaining active power P_r is carried across the air gap and transferred to the rotor by electromagnetic induction.

Due to the I^2R losses in the rotor, a third portion P_{jr} is dissipated as heat, and the remainder is finally available in the form of mechanical power P_m. By subtracting a small fourth portion P_v, representing windage and bearing-friction losses, we finally obtain P_L, the mechanical power available at the shaft to drive the load.

The power flow diagram of Fig. 13.15 enables us to identify and to calculate three important properties of the induction motor: (1) its *efficiency*, (2) its *power*, and (3) its *torque*.

1. Efficiency. By definition, the efficiency of a motor is the ratio of the output power to the input power:

$$\text{efficiency } (\eta) = P_L/P_e \qquad (13.6)$$

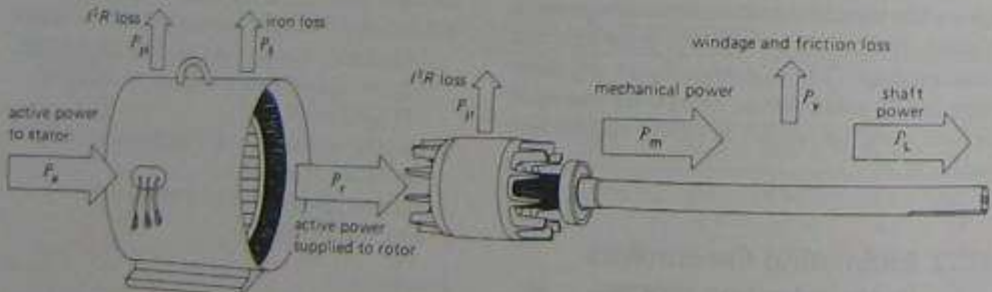


Figure 13.15
Active power flow in a 3-phase induction motor.

2. I^2R losses in the rotor. It can be shown* that the rotor I^2R losses P_{jr} are related to the rotor input power P_r by the equation

$$P_{jr} = sP_r \qquad (13.7)$$

where

P_{jr} = rotor I^2R losses [W]
s = slip
P_r = power transmitted to the rotor [W]

Equation 13.7 shows that as the slip increases, the rotor I^2R losses consume a larger and larger proportion of the power P_r transmitted across the air gap to the rotor. A rotor turning at half synchronous speed ($s = 0.5$) dissipates in the form of heat 50 percent of the active power it receives. When the rotor is locked ($s = 1$), all the power transmitted to the rotor is dissipated as heat.

3. Mechanical power. The mechanical power P_m developed by the motor is equal to the power transmitted to the rotor minus its I^2R losses. Thus,

$$\begin{aligned} P_m &= P_r - P_{jr} \\ &= P_r - sP_r \end{aligned} \qquad (13.7)$$

whence

$$P_m = (1 - s)P_r \qquad (13.8)$$

The actual mechanical power available to drive the load is slightly less than P_m, due to the power needed to overcome the windage and friction losses. In most calculations we can neglect this small loss.

4. Motor torque. The torque T_m developed by the motor at *any* speed is given by

$$T_m = \frac{9.55\, P_m}{n} \qquad (3.5)$$

$$= \frac{9.55\,(1 - s)\, P_r}{n_s(1 - s)} = 9.55\, P_r/n_s$$

therefore,

$$T_m = 9.55\, P_r/n_s \qquad (13.9)$$

where

T_m = torque developed by the motor at *any* speed [N·m]
P_r = power transmitted to the rotor [W]
n_s = synchronous speed [r/min]
9.55 = multiplier to take care of units [exact value: $60/2\pi$]

The actual torque T_L available at the shaft is slightly less than T_m, due to the torque required to overcome the windage and friction losses. However, in most calculations we can neglect this small difference.

Equation 13.9 shows that the torque is directly proportional to the active power transmitted to the rotor. Thus, to develop a high locked-rotor torque, the rotor must absorb a large amount of active power. The latter is dissipated in the form of heat, consequently, the temperature of the rotor rises very rapidly.

Example 13-5

A 3-phase induction motor having a synchronous speed of 1200 r/min draws 80 kW from a 3-phase

---

* $$\begin{bmatrix}\text{mechanical}\\ \text{power output}\\ \text{of rotor}\end{bmatrix} = \begin{bmatrix}\text{electromagnetic}\\ \text{power transferred}\\ \text{to rotor}\end{bmatrix} - \begin{bmatrix}\text{electrical}\\ \text{losses}\\ \text{in rotor}\end{bmatrix}$$

$$P_m = P_r - P_{jr} \qquad \text{(i)}$$

but from Eq. 3.5

$$P_m = \frac{\text{rotor speed} \times \text{mechanical torque}}{9.55}$$

Hence,

$$P_m = \frac{nT_m}{9.55} \qquad \text{(ii)}$$

Also from Eq. 3.5 we can write

$$P_r = \frac{\text{speed of flux} \times \text{electromagnetic torque}}{9.55}$$

$$P_r = \frac{n_s T_{mag}}{9.55} \qquad \text{(iii)}$$

but the mechanical torque T_m must equal the electromagnetic torque T_{mag}.

Thus

$$T_m = T_{mag} \qquad \text{(iv)}$$

Substituting (ii), (iii), and (iv) in (i), we find

$$P_{jr} = sP_r$$

feeder. The copper losses and iron losses in the stator amount to 5 kW. If the motor runs at 1152 r/min, calculate the following:

a. The active power transmitted to the rotor
b. The rotor I^2R losses
c. The mechanical power developed
d. The mechanical power delivered to the load, knowing that the windage and friction losses are equal to 2 kW
e. The efficiency of the motor

Solution

a. Active power to the rotor is

$$P_r = P_e - P_{js} - P_f$$
$$= 80 - 5 = 75 \text{ kW}$$

b. The slip is

$$s = (n_s - n)/n_s$$
$$= (1200 - 1152)/1200$$
$$= 48/1200 = 0.04$$

Rotor I^2R losses are

$$P_{jr} = sP_r = 0.04 \times 75 = 3 \text{ kW}$$

c. The mechanical power developed is

$$P_m = P_r - I^2R \text{ losses in rotor}$$
$$= 75 - 3 = 72 \text{ kW}$$

d. The mechanical power P_L delivered to the load is slightly less than P_m, due to the friction and windage losses.

$$P_L = P_m - P_v = 72 - 2 = 70 \text{ kW}$$

e. The efficiency is

$$\eta = P_L/P_e = 70/80$$
$$= 0.875 \text{ or } 87.5\%$$

Example 13-6

A 3-phase, 8-pole squirrel-cage induction motor, connected to a 60 Hz line, possesses a synchronous speed of 900 r/min. The motor absorbs 40 kW, and the copper and iron losses in the stator amount to 5 kW and 1 kW, respectively. Calculate the torque developed by the motor.

Solution

The power transmitted across the air gap to the rotor is

$$P_r = P_e - P_{js} - P_f$$
$$= 40 - 5 - 1 = 34 \text{ kW}$$

$$T_m = 9.55\, P_r/n_s \qquad (13.9)$$
$$= 9.55 \times 34\,000/900$$
$$= 361 \text{ N·m}$$

Note that the solution to this problem (the torque) is independent of the speed of rotation. The motor could be at a standstill or running at full speed, but as long as the power P_r transmitted to the rotor is equal to 34 kW, the motor develops a torque of 361 N·m.

Learning Outcome 1.3 Continue

Winding Calculation

&

Learning Outcome 1..7

Perform calculations using the relationship between air gap power, losses and net torque

Figure 13.22a
The 24 coils are grouped two-by-two to make 12 groups.

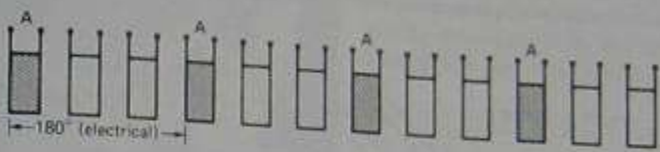


Figure 13.22b
The four groups of phase A are selected so as to be evenly spaced from each other.

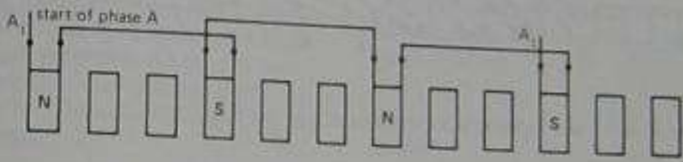


Figure 13.22c
The groups of phase A are connected in series to create alternate N-S poles.

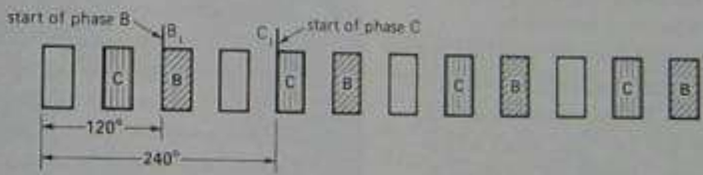


Figure 13.22d
The start of phases B and C begins 120° and 240°, respectively, after the start of phase A.

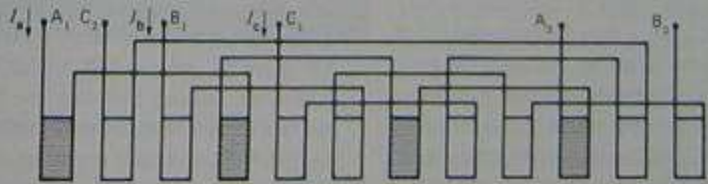


Figure 13.22e
When all phase groups are connected, only six leads remain.

coil side set in each slot. If the windings are now laid down so that all the other coil sides fall into the slots, we obtain the classical appearance of a 3-phase lap winding having two coil sides per slot (Fig. 13.21b). The coils are connected together to create three identical windings, one for each phase. Each winding consists of a number of groups equal to the number of poles. The groups of each phase are symmetrically distributed around the circumference of the stator. The following examples show how this is done.

Example 13-8

The stator of a 3-phase, 10-pole induction motor possesses 120 slots. If a lap winding is used, calculate the following:

a. The total number of coils
b. The number of coils per phase
c. The number of coils per group
d. The pole pitch
e. The coil pitch (expressed as a percentage of the pole pitch), if the coil width extends from slot 1 to slot 11

Solution

a. A 120-slot stator requires 120 coils.
b. Coils per phase = 120 ÷ 3 = 40.

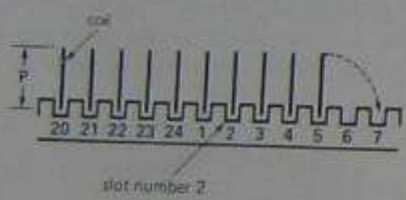


Figure 13.21a
Coils held upright in 24 stator slots.

Figure 13.21b
Coils laid down to make a typical lap winding.

c. Number of groups per phase = number of poles = 10
   Coils per group = 40 ÷ 10 = 4.
d. The pole pitch corresponds to
   pole pitch = slots/poles = 120/10
   = 12 slots
   One pole pitch extends therefore from slot 1 (say) to slot 13.
e. The coil pitch covers 10 slots (slot 1 to slot 11).
   The percent coil pitch = 10/12 = 83.3%.

The next example shows in greater detail how the coils are interconnected in a typical 3-phase stator winding.

Example 13-9

A stator having 24 slots has to be wound with a 3-phase, 4-pole winding. Determine the following:

1. The connections between the coils
2. The connections between the phases

Solution

The 3-phase winding has 24 coils. Assume that they are standing upright, with one coil side in each slot (Fig. 13.22). We will first determine the coil distribution for phase A and then proceed with the connections for that phase. Similar connections will then be made for phases B and C. Here is the line of reasoning:

a. The revolving field creates 4 poles; the motor therefore has 4 groups per phase, or 4 × 3 = 12 phase groups in all. Each rectangle in Fig. 13.22a represents one group. Because the stator contains 24 coils, each group consists of 24/12 = 2 consecutive coils.
b. The groups (poles) of each phase must be uniformly spaced around the stator. The group distribution for phase A is shown in Fig. 13.22b. Each shaded rectangle represents two upright coils connected in series, producing the two terminals shown. Note that the mechanical distance between two successive groups always corresponds to an *electrical* phase angle of 180°.
c. Successive groups of phase A must have opposite magnetic polarities. Consequently, the four

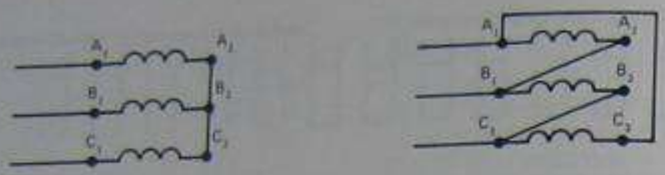


Figure 13.22f
The phase may be connected in wye or in delta, and three leads are brought out to the terminal box.

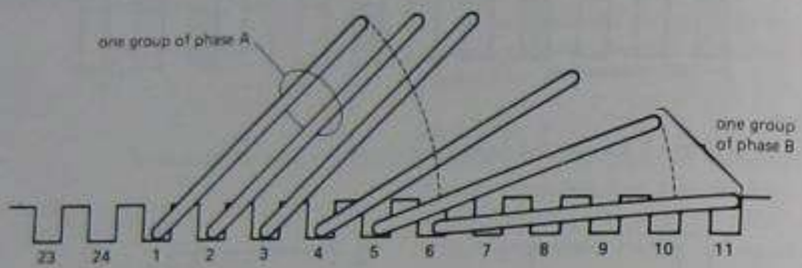


Figure 13.23
The pole pitch is from slot 1 slot 7; the coil pitch from slot 1 to slot 6.

groups of phase A are connected in series to produce successive N-S-N-S poles (Fig. 13.22c). Phase A now has two terminals, a *starting* terminal A_1 and a *finishing* terminal A_2.

d. The phase groups of phases B and C are spaced the same way around the stator. However, the *starting* terminals B_1 and C_1 are respectively located at 120° and 240° (electrical) with respect to the starting terminal A_1 of phase A (Fig. 13.22d).

e. The groups in phases B and C are connected in series in the same way as those of phase A are (Fig. 13.22e). This yields six terminals: A_1A_2, B_1B_2, and C_1C_2. They may be connected either in wye or in delta inside the machine. The resulting 3 wires corresponding to the 3 phases are brought out to the terminal box of the machine (Fig. 13.22f). In practice, the connections are made, not while the coils are upright (as shown) but only after they have been laid down in the slots.

f. Because the pole pitch corresponds to a span of $24/4 = 6$ slots, the coil pitch may be shortened to 5 slots (slot 1 to slot 6). Thus, the first coil of phase A is lodged in the first and sixth slots (Fig. 13.23). All the other coils and connections follow suit according to Fig. 13.22e.

Figs. 13.24a and 13.24b show the coil and stator of a 450 kW (600 hp) induction motor. Fig. 13.25 illustrates the procedure used in winding a smaller 37.5 kW (50 hp) stator.

13.18 Sector motor

Consider a standard 3-phase, 4-pole, wye-connected motor having a synchronous speed of 1800 r/min. Let us cut the stator in half, so that half the winding is removed and only two complete N and S poles are left (per phase). Next, let us connect the three phases in wye, without making any other changes to the existing coil connections. Finally, we mount the original rotor above this *sector stator*, leaving a small air gap (Fig. 13.26).

Figure 13.24a
Stator of a 3-phase, 450 kW, 1180 r/min, 575 V, 60 Hz induction motor. The lap winding is composed of 108 preformed coils having a pitch from slots 1 to 15. One coil side falls into the bottom of a slot and the other at the top. Rotor diameter: 500 mm; axial length: 460 mm. *(Courtesy of Services Électro-mécaniques Roberge)*

Figure 13.24b
Close-up view of the preformed coil in Fig. 13.24a.

If we connect the stator terminals to a 3-phase, 60 Hz source, the rotor will again turn at close to 1800 r/min. To prevent saturation, the voltage should be reduced to half its original value because the stator winding now has only one-half the original number of turns. Under these conditions, this remarkable truncated *sector motor* still develops about 20 percent of its original rated power.

The sector motor produces a *revolving* field that moves at the same peripheral speed as the flux in the original 3-phase motor. However, instead of making a complete turn, the field simply travels continuously from one end of the stator to the other.

13.19 Linear induction motor

It is obvious that the sector stator could be laid out flat, without affecting the shape or speed of the magnetic field. Such a flat stator produces a field that moves at constant speed, in a straight line. Using the same reasoning as in Section 13.5, we can prove that the flux travels at a linear synchronous speed given by

$$v_s = 2wf \tag{13.10}$$

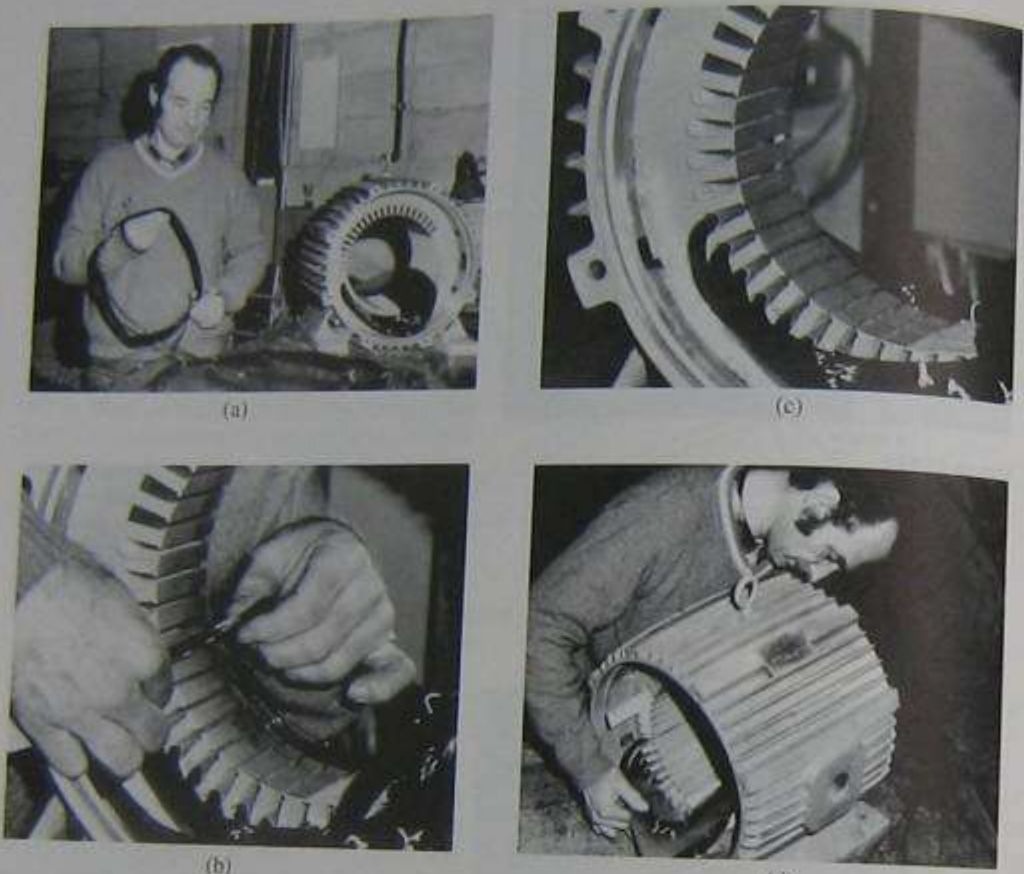
where

v_s = linear synchronous speed [m/s]
w = width of one pole-pitch [m]
f = frequency [Hz]

Note that the linear speed does not depend upon the number of poles but only on the pole-pitch. Thus, it is possible for a 2-pole linear stator to create a field moving at the same speed as that of a 6-pole linear stator (say), provided they have the same pole-pitch.

If a flat squirrel-cage winding is brought near the flat stator, the travelling field drags the squirrel cage along with it (Section 13.2). In practice, we generally use a simple aluminum or copper plate as a rotor (Fig. 13.27). Furthermore, to increase the power and to reduce the reluctance of the magnetic path, two flat stators are usually mounted, face-to-face, on opposite sides of the aluminum plate. The combination is called a *linear induction motor*. The direction of the motor can be reversed by interchanging any two stator leads.

In many practical applications, the rotor is stationary while the stator moves. For example, in some high-speed trains, the rotor is composed of a



(a) (c) (b) (d)

Figure 13.25
Stator winding of a 3-phase, 50 hp, 575 V, 60 Hz, 1764 r/min induction motor. The stator possesses 48 slots carrying 48 coils connected in wye.

a. Each coil is composed of 5 turns of five No. 15 copper wires connected in parallel. The wires are covered with a high-temperature polyimide insulation. Five No. 15 wires in parallel is equivalent to one No. 8 wire.
b. One coil side is threaded into slot 1 (say) and the other side goes into slot 12. The coil pitch is, therefore, from 1 to 12.
c. Each coil side fills half a slot and is covered with a paper spacer so that it does not touch the second coil side placed in the same slot. Starting from the top, the photograph shows 3 empty and uninsulated slots and 4 empty slots insulated with a composition paper liner. The remaining 10 slots each carry one coil side.
d. A varnished cambric cloth, cut in the shape of a triangle, provides extra insulation between adjacent phase groups.

(Courtesy of Services Électromécaniques Roberge)

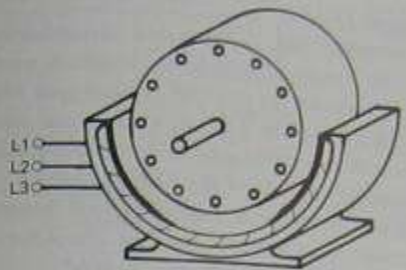


Figure 13.26
Two-pole sector induction motor.

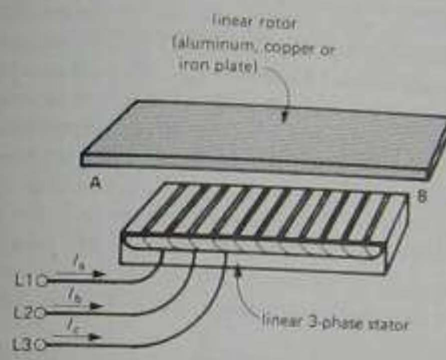


Figure 13.27
Components of a 3-phase linear induction motor.

thick aluminum plate fixed to the ground and extending over the full length of the track. The linear stator is bolted to the undercarriage of the train and straddles the plate. Train speed is varied by changing the frequency applied to the stator (Fig. 13.31).

Example 13-10

The stator of a linear induction motor is excited from a 75 Hz electronic source. If the distance between consecutive phase groups of phase A is 300 mm, calculate the linear speed of the magnetic field.

Solution

The pole pitch is 300 mm. Consequently,

$$v_s = 2\,wf \qquad (13.10)$$
$$= 2 \times 0.3 \times 75$$
$$= 45 \text{ m/s or } 162 \text{ km/h}$$

13.20 Traveling waves

We are sometimes left with the impression that when the flux reaches the end of a linear stator, there must be a delay before it returns to restart once more at the beginning. This is not the case. The linear motor produces a traveling wave of flux which moves continuously and smoothly from one end of the stator to the other. Figure 13.28 shows how the flux moves from left to right in a 2-pole linear motor. The flux cuts off sharply at extremities A, B of the stator. However, as fast as a N or S pole disappears at the right, it builds up again at the left.

13.21 Properties of a linear induction motor

The properties of a linear induction motor are almost identical to those of a standard rotating machine. Consequently, the equations for slip, thrust, power, etc., are also similar.

1. Slip. Slip is defined by

$$s = (v_s - v)/v_s \qquad (13.11)$$

where

s = slip
v_s = synchronous linear speed [m/s]
v = speed of rotor (or stator) [m/s]

2. Active power flow. With reference to Fig. 13.15, active power flows through a linear motor in the same way it does through a rotating motor, except that the stator and rotor are flat. Consequently, Eqs. 13.6, 13.7, and 13.8 apply to both types of machines:

$$\eta = P_L/P_e \qquad (13.6)$$
$$P_{jr} = sP_r \qquad (13.7)$$
$$P_m = (1 - s)P_r \qquad (13.8)$$

3. Thrust. The thrust or force developed by a linear induction motor is given by:

$$F = P_r/v_s \qquad (13.12)$$

where

F = thrust [N]
P_r = power transmitted to the rotor [W]
v_s = linear synchronous speed [m/s]

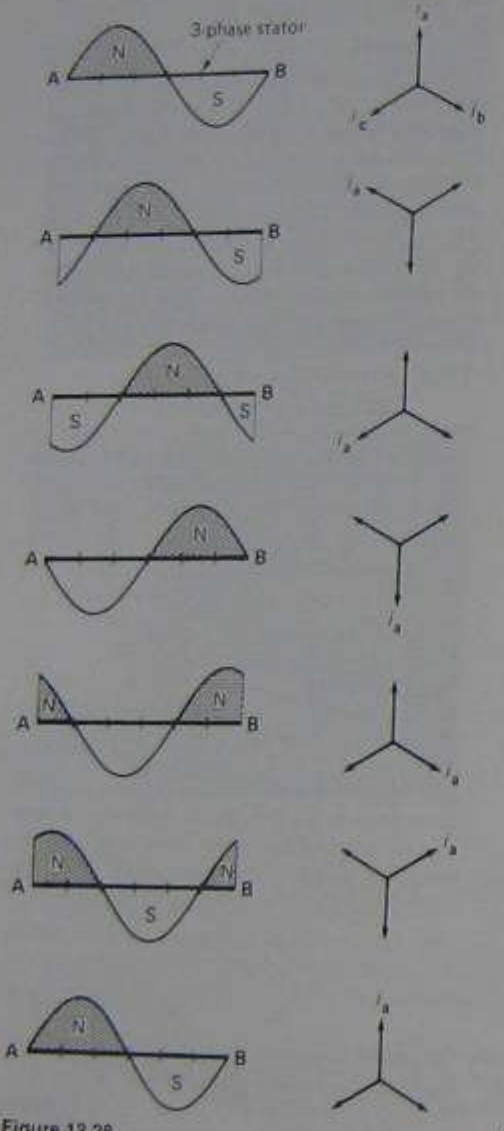


Figure 13.28
Shape of the magnetic field created by a 2-pole, 3-phase linear stator, over one complete cycle. The successive frames are separated by an interval of time equal to 1/6 cycle or 60°.

Example 13-11

An overhead crane in a factory is driven horizontally by means of two linear induction motors whose rotors are the two steel I-beams upon which the crane rolls. The 3-phase, 4-pole linear stators (mounted on opposite sides of the crane and facing the respective webs of the I-beams) have a pole pitch of 8 cm and are driven by a variable frequency electronic source. During a test on one of the motors, the following results were obtained:

stator frequency: 15 Hz
power to stator: 5 kW
copper loss + iron loss in stator: 1 kW
crane speed: 1.8 m/s

Calculate

a. Synchronous speed and slip
b. Power to the rotor
c. I^2R loss in rotor
d. Mechanical power and thrust

Solution

a. Linear synchronous speed

$$\begin{aligned} v_s &= 2\,wf \qquad (13.10) \\ &= 2 \times 0.08 \times 15 \\ &= 2.4 \text{ m/s} \end{aligned}$$

The slip is

$$\begin{aligned} s &= (v_s - v)/v_s \qquad (13.11) \\ &= (2.4 - 1.8)/2.4 \\ &= 0.25 \end{aligned}$$

b. Power to the rotor is

$$\begin{aligned} P_r &= P_e - P_{js} - P_f \qquad \text{(see Fig. 13.15)} \\ &= 5 - 1 \\ &= 4 \text{ kW} \end{aligned}$$

c. I^2R loss in the rotor is

$$\begin{aligned} P_{jr} &= sP_r \qquad (13.7) \\ &= 0.25 \times 4 \\ &= 1 \text{ kW} \end{aligned}$$

d. Mechanical power is

$$\begin{aligned} P_m &= P_r - P_{jr} \quad \text{(Fig. 13.15)} \\ &= 4 - 1 \\ &= 3 \text{ kW} \end{aligned}$$

The thrust is

$$\begin{aligned} F &= P_r/v_s \quad (13.12) \\ &= 4000/2.4 \\ &= 1667 \text{ N} = 1.67 \text{ kN } (\sim 375 \text{ lb}) \end{aligned}$$

13.22 Magnetic levitation

In Section 13.2 we saw that a moving permanent magnet, sweeping across a conducting ladder, tends to drag the ladder along with the magnet. We will now show that this horizontal tractive force is also accompanied by a *vertical* force, which tends to push the magnet away from the ladder.

Referring to Fig. 13.29, suppose that conductors 1, 2, 3 are three conductors of the stationary ladder. The center of the N pole of the magnet is sweeping across the top of conductor 2. The voltage induced in this conductor is maximum because the flux density is greatest at the center of the pole. If the magnet moves very slowly, the resulting induced current reaches *its* maximum value at virtually the same time. This current, returning by conductors 1 and 3, creates magnetic poles nnn and sss as shown in Fig. 13.29. According to the laws of attraction and repulsion, the front half of the magnet is repelled upward while the rear half is attracted downward. Because the distribution of the nnn and sss poles is symmetrical with respect to the center of the magnet, the vertical forces of attraction and repulsion are equal, and the resulting vertical force is nil. Consequently, there is only a horizontal tractive force.

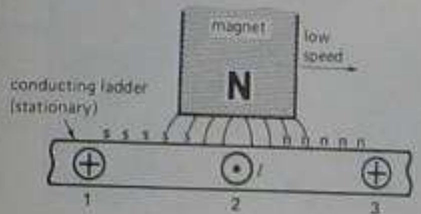


Figure 13.29
Currents and magnetic poles at low speed.

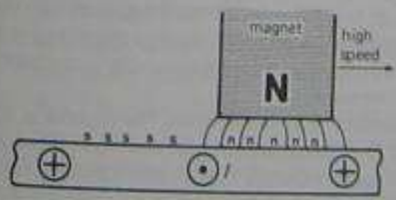


Figure 13.30
Currents and magnetic poles at high speed.

But suppose now that the magnet moves very rapidly. Owing to its inductance, the current in conductor 2 reaches its maximum value a fraction of a second after the voltage has attained *its* maximum. Consequently, by the time the current in conductor 2 is maximum, the center of the magnet is already some distance ahead of the conductor (Fig. 13.30). The current returning by conductors 1 and 3 again creates nnn and sss poles; however, the N pole of the magnet is now directly above an nnn pole, with the result that a strong vertical force tends to push the magnet upward.* This effect is called the principle of *magnetic levitation*.

Magnetic levitation is used in some ultra-high-speed trains that glide on a magnetic cushion rather than on wheels. A powerful electromagnet fixed underneath the train moves above a conducting rail inducing currents in the rail in the same way as in our ladder. The force of levitation is always accompanied by a small horizontal braking force which must, of course, be overcome by the linear motor that propels the train. See Figs. 13.31 and 13.32.

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* The current is always delayed (even at low speeds) by an interval of time Δt, which depends upon the L/R time constant of the rotor. This delay is so brief that, at low speeds, the current reaches its maximum at virtually the same time and place as the voltage does. On the other hand, at high speeds, the same delay Δt produces a significant shift in space between the points where the voltage and current reach their respective maximum values.

ELECTROMAGNETISM

2.16 Magnetic field intensity *H* and flux density *B*

Whenever a magnetic flux ϕ exists in a body or component, it is due to the presence of a magnetic field *intensity* H, given by

$$H = U/l \qquad (2.18)$$

where

H = magnetic field intensity [A/m]
U = magnetomotive force acting on the component [A] (or ampere turn)
l = length of the component [m]

The resulting magnetic flux density is given by

$$B = \phi/A \qquad (2.19)$$

where

B = flux density [T]
ϕ = flux in the component [Wb]
A = cross section of the component [m^2]

There is a definite relationship between the flux density (B) and the magnetic field intensity (H) of any material. This relationship is usually expressed graphically by the B-H curve of the material.

2.17 *B-H* curve of vacuum

In vacuum, the magnetic flux density B is directly proportional to the magnetic field intensity H, and is expressed by the equation

$$B = \mu_o H \qquad (2.20)$$

where

B = flux density [T]
H = magnetic field intensity [A/m]
μ_o = magnetic constant [$= 4\pi \times 10^{-7}$]*

* Also called the permeability of vacuum. The complete expression for μ_o is $4\pi \times 10^{-7}$ henry/meter.

In the SI, the magnetic constant is fixed, by definition. It has a numerical value of $4\pi \times 10^{-7}$ or approximately 1/800 000. This enables us to write Eq. 2-20 in the approximate form:

$$H = 800\,000\,B \qquad (2.21)$$

The B-H curve of vacuum is a straight line. A vacuum never saturates, no matter how great the flux density may be (Fig. 2.25). The curve shows that a magnetic field intensity of 800 A/m produces a flux density of 1 millitesla.

Nonmagnetic materials such as copper, paper, rubber, and air have B-H curves almost identical to that of vacuum.

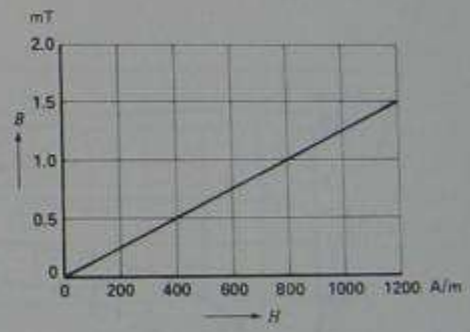


Figure 2.25
B-H curve of vacuum and of nonmagnetic materials.

2.18 *B-H* curve of a magnetic material

The flux density in a magnetic material also depends upon the magnetic field intensity to which it is subjected. Its value is given by

$$B = \mu_o \mu_r H \qquad (2.22)$$

where B, μ_o, and H have the same significance as before, and μ_r is the relative permeability of the material.

The value of μ_r is not constant but varies with the flux density in the material. Consequently, the relationship between B and H is not linear, and this makes Eq. 2.22 rather impractical to use. We prefer to show the relationship by means of a B-H saturation curve. Thus, Fig. 2.26 shows typical saturation curves of three materials commonly used in electrical machines: silicon iron, cast iron, and cast steel. The curves show that a magnetic field intensity of 2000 A/m produces a flux density of 1.4 T in cast steel but only 0.5 T in cast iron.

2.19 Determining the relative permeability

The *relative permeability* μ_r of a material is the ratio of the flux density in the material to the flux density that would be produced in vacuum, under the same magnetic field intensity H.

Given the saturation curve of a magnetic material, it is easy to calculate the relative permeability using the approximate equation

$$\mu_r \approx 800\,000\ B/H \tag{2.23}$$

where

B = flux density in the magnetic material [T]

H = corresponding magnetic field intensity [A/m]

Example 2-7

Determine the permeability of silicon iron (1%) at a flux density of 1.4 T.

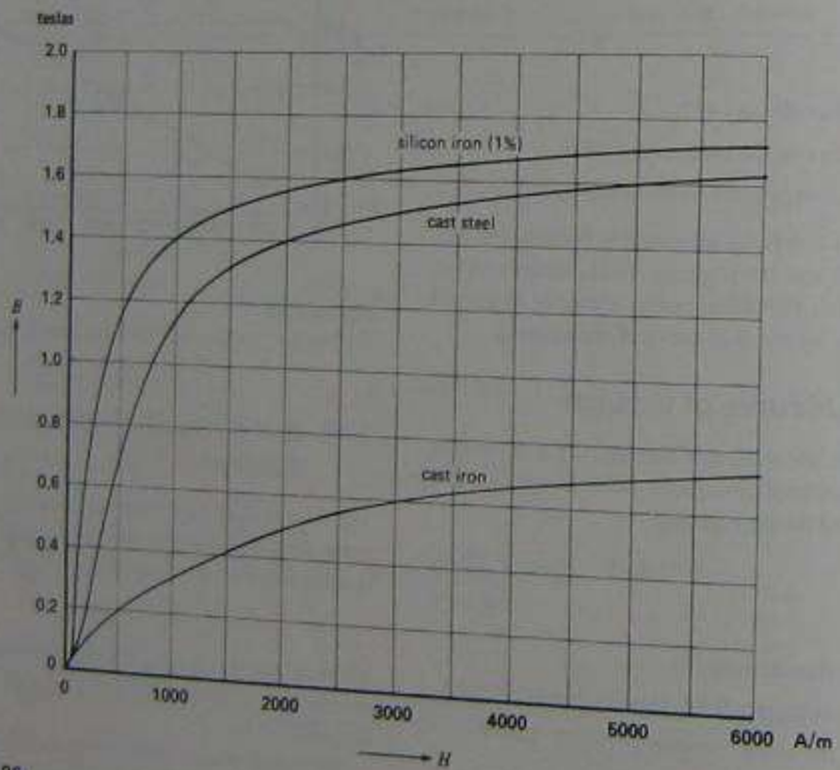


Figure 2.26

B-H saturation curves of three magnetic materials.

Solution

Referring to the saturation curve (Fig. 2.26), we see that a flux density of 1.4 T requires a magnetic field intensity of 1000 A/m. Consequently,

$$\mu_r = 800\,000\ B/H$$
$$= 800\,000 \times 1.4/1000 = 1120$$

At this flux density, silicon iron is 1120 times more permeable than vacuum (or air).

Fig. 2.27 shows the saturation curves of a broad range of materials from vacuum to Permalloy®, one of the most permeable magnetic materials known. Note that as the magnetic field intensity increases, the magnetic materials saturate more and more and eventually all the *B-H* curves follow the *B-H* curve of vacuum.

2.20 Faraday's law of electromagnetic induction

In 1831, while pursuing his experiments, Michael Faraday made one of the most important discoveries in electromagnetism. Now known as **Faraday's law of electromagnetic induction**, it revealed a fundamental relationship between the voltage and flux in a circuit. Faraday's law states:

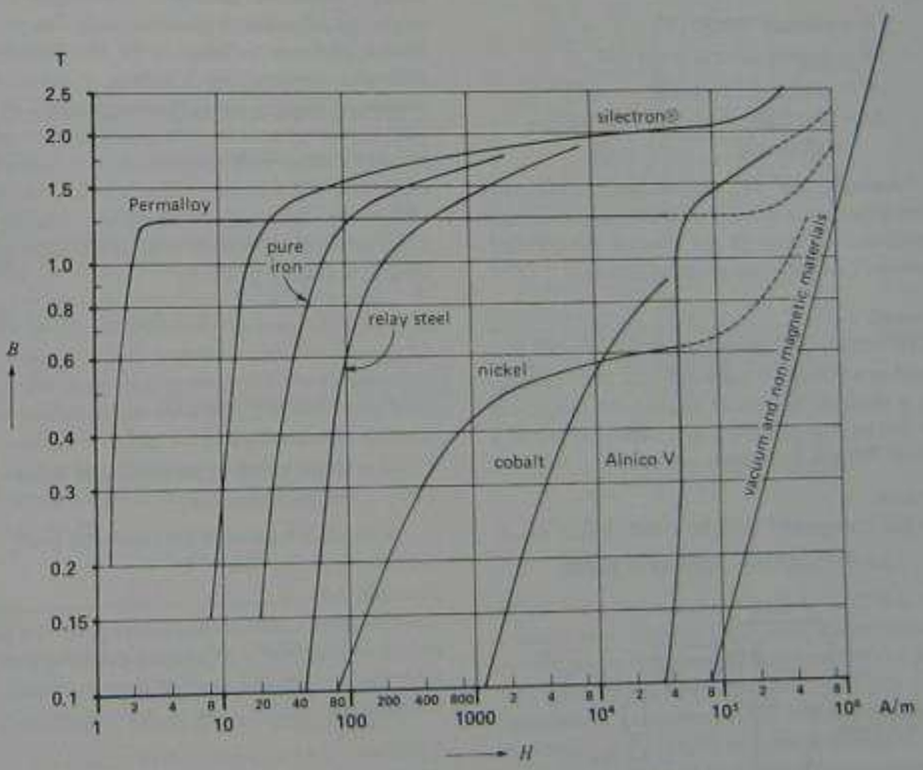


Figure 2.27
Saturation curves of magnetic and nonmagnetic materials. Note that all curves become asymptotic to the *B-H* curve of vacuum where *H* is high.

1. **If the flux linking a loop (or turn) varies as a function of time, a voltage is induced between its terminals.**
2. **The value of the induced voltage is proportional to the rate of change of flux.**

By definition, and according to the SI system of units, when the flux inside a loop varies at the rate of 1 weber per second, a voltage of 1 V is induced between its terminals. Consequently, if the flux varies inside a coil of N turns, the voltage induced is given by

$$E = N\frac{\Delta\Phi}{\Delta t} \tag{2.24}$$

where

E = induced voltage [V]
N = number of turns in the coil
$\Delta\Phi$ = change of flux inside the coil [Wb]
Δt = time interval during which the flux changes [s]

Faraday's law of electromagnetic induction opened the door to a host of practical applications and established the basis of operation of transformers, generators, and alternating current motors.

Example 2-8

A coil of 2000 turns surrounds a flux of 5 mWb produced by a permanent magnet (Fig. 2.28). The magnet is suddenly withdrawn causing the flux inside the coil to drop uniformly to 2 mWb in 1/10 of a second. What is the voltage induced?

Solution

The flux change is

$$\Delta\Phi = (5\ \text{mWb} - 2\ \text{mWb}) = 3\ \text{mWb}$$

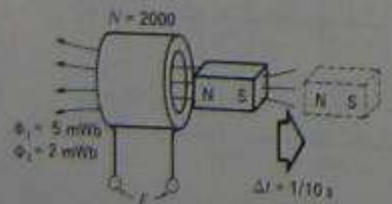


Figure 2.28
Voltage induced by a moving magnet. See Example 2-8.

Because this change takes place uniformly in 1/10 of a second (Δt), the induced voltage is

$$E = N\frac{\Delta\Phi}{\Delta t} = 2000 \times \frac{3}{1000 \times 1/10}$$

$$= 60\ \text{V}$$

The induced voltage falls to zero as soon as the flux ceases to change.

2.21 Voltage induced in a conductor

In many motors and generators, the coils move with respect to a flux that is fixed in space. The relative motion produces a change in the flux linking the coils and, consequently, a voltage is induced according to Faraday's law. However, in this special (although common) case, it is easier to calculate the induced voltage with reference to the *conductors*, rather than with reference to the coil itself. In effect, whenever a conductor cuts a magnetic field, a voltage is induced across its terminals. The value of the induced voltage is given by

$$E = Blv \tag{2.25}$$

where

E = induced voltage [V]
B = flux density [T]
l = active length of the conductor in the magnetic field [m]
v = relative speed of the conductor [m/s]

Example 2-9

The stationary conductors of a large generator have an active length of 2 m and are cut by a field of 0.6 teslas, moving at a speed of 100 m/s (Fig. 2.29). Calculate the voltage induced in each conductor.

Solution

According to Eq. 2-25, we find

$$E = Blv$$

$$= 0.6 \times 2 \times 100 = 120\ \text{V}$$

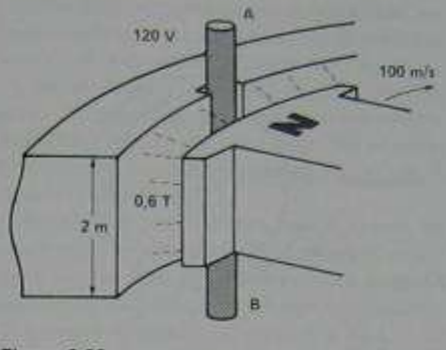


Figure 2.29
Voltage induced in a stationary conductor. See Example 2-9.

2.22 Lorentz force on a conductor

When a current-carrying conductor is placed in a magnetic field, it is subjected to a force which we call *electromagnetic force*, or Lorentz force. This force is of fundamental importance because it constitutes the basis of operation of motors, of generators, and of many electrical instruments. The magnitude of the force depends upon the orientation of the conductor with respect to the direction of the field. The force is greatest when the conductor is perpendicular to the field (Fig. 2.30) and zero when it is parallel to it (Fig. 2.31). Between these two extremes, the force has intermediate values.

The maximum force acting on a straight conductor is given by

$$F = BlI \qquad (2.26)$$

where

F = force acting on the conductor [N]
B = flux density of the field [T]
l = active length of the conductor [m]
I = current in the conductor [A]

Example 2-10

A conductor 3 m long carrying a current of 200 A is placed in a magnetic field whose density is 0.5 T. Calculate the force on the conductor if it is perpendicular to the lines of force (Fig. 2.30).

Solution

$$F = BlI$$
$$= 0.5 \times 3 \times 200 = 300 \text{ N}$$

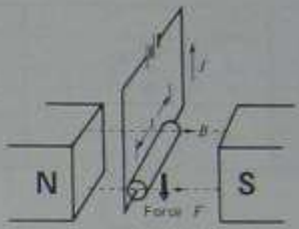


Figure 2.30
Force on a conductor.

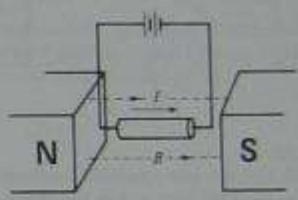


Figure 2.31
Force = 0.

2.23 Direction of the force acting on a straight conductor

Whenever a conductor carries a current, it is surrounded by a magnetic field. For a current flowing into the page of this book, the circular lines of force have the direction shown in Figure 2.32a. The same figure shows the magnetic field created between the N, S poles of a powerful permanent magnet.

The magnetic field does not, of course, have the shape shown in the figure because lines of force never cross each other. What, then, is the shape of the resulting field?

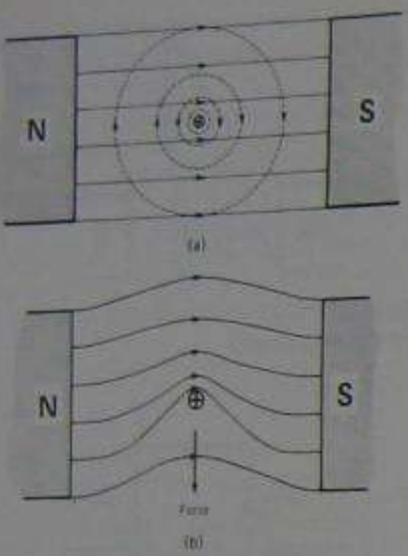


Figure 2.32
a. Magnetic field due to magnet and conductor.
b. Resulting magnetic field pushes the conductor downward.

To answer the question, we observe that the lines of force created respectively by the conductor and the permanent magnet act in the same direction above the conductor and in opposite directions below it. Consequently, the number of lines above the conductor must be greater than the number below. The resulting magnetic field therefore has the shape given in Figure 2.32b.

Recalling that lines of flux act like stretched elastic bands, it is easy to visualize that a force acts upon the conductor, tending to push it downward.

2.24 Residual flux density and coercive force

Consider the coil of Figure 2.33a, which surrounds a magnetic material formed in the shape of a ring. A current source, connected to the coil, produces a current whose value and direction can be changed at will. Starting from zero, we gradually increase I, so that H and B increase. This increase traces out

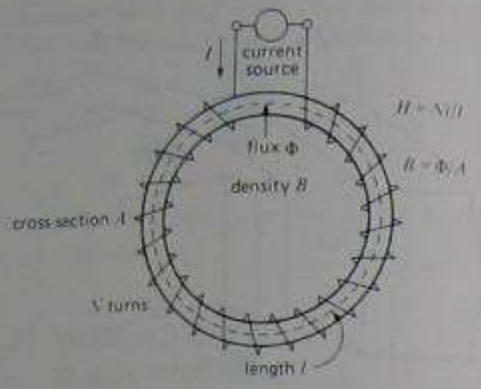


Figure 2.33a
Method of determining the *B-H* properties of a magnetic material.

curve **oa** in Figure 2.33b. The flux density reaches a value B_m for a magnetic field intensity H_m.

If the current is now gradually reduced to zero, the flux density B does not follow the original curve, but moves along a curve **ab** situated above **oa**. In effect, as we reduce the magnetic field intensity, the magnetic domains that were lined up under the influence of field H_m tend to retain their original orientation. This phenomenon is called hysteresis. Consequently, when H is reduced to zero, a substantial flux density remains. It is called residual flux density or *residual induction* (B_r).

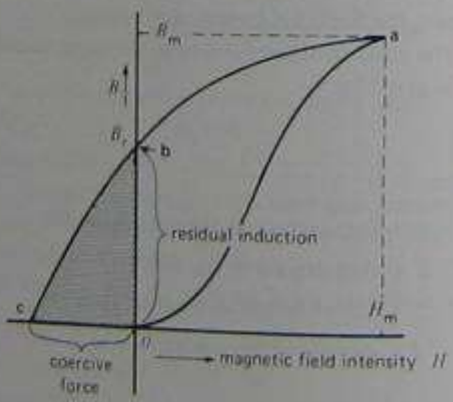


Figure 2.33b
Residual induction and coercive force.

If we wish to eliminate this residual flux, we have to reverse the current in the coil and gradually increase H in the opposite direction. As we do so, we move along curve **bc.** The magnetic domains gradually change their previous orientation until the flux density becomes zero at point **c.** The magnetic field intensity required to reduce the flux to zero is called *coercive force* (H_c).

In reducing the flux density from B_r to zero, we also have to furnish energy. This energy is used to overcome the frictional resistance of the magnetic domains as they oppose the change in orientation. The energy supplied is dissipated as heat in the material. A very sensitive thermometer would indicate a slight temperature rise as the ring is being demagnetized.

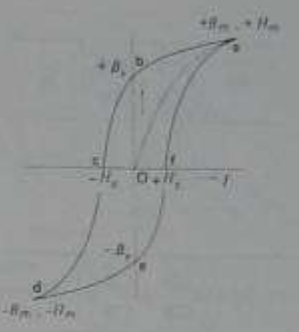


Figure 2.34
Hysteresis loop. If B is expressed in teslas and H in amperes per meter, the area of the loop is the energy dissipated per cycle, in joules per kilogram.

2.25 Hysteresis loop

Transformers and most electric motors operate on alternating current. In such devices the flux in the iron changes continuously both in value and direction. The magnetic domains are therefore oriented first in one direction, then the other, at a rate that depends upon the frequency. Thus, if the flux has a frequency of 60 Hz, the domains describe a complete cycle every 1/60 of a second, passing successively through peak flux densities $+B_m$ and $-B_m$ as the peak magnetic field intensity alternates between $+H_m$ and $-H_m$. If we plot the flux density B as a function of H, we obtain a closed curve called hysteresis loop (Fig. 2.34). The residual induction B_r and coercive force H_c have the same significance as before.

2.26 Hysteresis loss

In describing a hysteresis loop, the flux moves successively from $+B_m$, $+B_r$, 0, $-B_m$, $-B_r$, 0, and $+B_m$, corresponding respectively to points **a, b, c, d, e, f,** and **a,** of Figure 2.34. The magnetic material absorbs energy during each cycle and this energy is dissipated as heat. We can prove that the amount of heat released per cycle (expressed in J/m³) is equal to the area (in T·A/m) of the hysteresis loop.

To reduce hysteresis losses, we select magnetic materials that have a narrow hysteresis loop, such as the grain-oriented silicon steel used in the cores of alternating-current transformers.

2.27 Hysteresis losses caused by rotation

Hysteresis losses are also produced when a piece of iron rotates in a constant magnetic field. Consider, for example, an armature AB, made of iron, that revolves in a field produced by permanent magnets N, S (Fig. 2.35). The magnetic domains in the armature tend to line up with the magnetic field, irrespective of the position of the armature. Consequently, as the armature rotates, the N poles of the domains point first toward A and then toward B. A complete reversal occurs therefore every half-revolution, as can be seen in Fig. 2.35a and 2.35b. Consequently, the magnetic domains in the armature reverse periodically, even though the magnetic field is constant. Hysteresis losses are produced just as they are in an ac magnetic field.

2.4 Verify the predicted motor performance at a given condition of load (Background theory for practical test)

INDUCTION MACHINES

It was shown in Chapter 11 that, when a polyphase stator winding is excited from a balanced polyphase supply, a stator m.m.f. distribution is set up and travels at synchronous speed given by eqn. (11.10) as

$$n_0 = \frac{f}{p} \tag{13.1}$$

Associated with the stator m.m.f. distribution is a flux density distribution which also travels at synchronous speed and is often referred to as a "rotating field".

The stator field induces voltages in the rotor phase windings so that a rotor m.m.f. distribution and an associated flux density distribution are set up. The rotor distributions travel at the same speed as the stator distribution. The axes of the stator and rotor distributions have an angular displacement, and as a result a torque acts on the rotor and causes it to accelerate in the same direction as the stator field.

The steady-state rotor speed is normally slightly less than synchronous so that the motor runs with a *per-unit slip*, s, defined as

$$s = \frac{n_0 - n_r}{n_0} \tag{13.2}$$

where n_r is the rotor speed.

At standstill, $n_r = 0$ and $s = 1$. For the rotor to reach synchronous speed ($n_r = n_0$ and $s = 0$), an external drive is necessary, since for this condition there is no rotor e.m.f. and hence no rotor current or torque. If the rotor is driven so that $n_r > n_0$, the slip becomes negative, the rotor torque opposes the external driving torque and the machine acts as an induction generator.

In all cases the slip speed is

$$n_s = n_0 - n_r \tag{13.3}$$

From eqn. (13.2),

$$n_s = sn_0 \tag{13.4}$$

and

$$n_r = (1 - s)n_0 \tag{13.5}$$

The frequency of the rotor e.m.f.s and currents is proportional to the difference in speed between the rotating field and the rotor, so that

$$f_r = (n_0 - n_r)p = sn_0p = sf$$

where p is the number of pole pairs. Hence

$$\frac{f_r}{f} = s \tag{13.6}$$

Similarly for the angular frequencies corresponding to f_r and f:

$$\frac{\omega_r}{\omega_0} = \frac{2\pi f_r}{2\pi f} = s \tag{13.7}$$

The 3-phase induction motor has a torque characteristic similar to that of the d.c. shunt motor, is robust, and is low in initial cost. Other forms of asynchronous machine are the a.c. commutator motor, which gives a wide range of speed control, and various types of single-phase motor, which are employed for fractional-horsepower drives, in individual units, and in traction.

13.1 Construction

The induction machine consists essentially of a stator, which carries a 3-phase winding, and a rotor. The stator winding is a 3-phase winding of one of the types described in Chapter 11, often being a narrow-spread mesh-connected closed winding. The winding is laid in open or half-closed slots in a laminated silicon-steel core.

The rotor winding is placed in half-closed or closed slots, the air-gap between stator and rotor being reduced to a minimum. There are two main types of rotor, the *wound rotor* and the *squirrel-cage rotor*. In the squirrel-cage rotor, solid conducting rods are inserted into closed slots, and at each end the rods are connected to a heavy short-circuiting ring. This forms a permanently short-circuited winding which is practically indestructible. In some smaller machines the conductors, end rings and fan are cast in one piece in aluminium. The cage rotor is cheap and robust, but suffers from the disadvantage of a low starting torque.

The wound rotor has a 3-phase winding with the same number of poles as the stator; the ends of the rotor winding may be brought out to three slip rings. The advantage of the wound-rotor machine is that an external starting resistance can be connected to the slip rings to give a large starting torque. This resistance is reduced to zero as the machine runs up to speed.

13.2 Equivalent Circuit of Induction Machine at Any Slip

The approximate equivalent circuit per phase of a polyphase induction machine at standstill ($s = 1$) is shown in Fig. 13.1. The equivalent circuit takes the same form as that adopted for the power transformer, since at standstill the induction machine consists of two polyphase windings linked by a common flux.

Unlike that of a power transformer, the magnetic circuit of the induction machine has an air-gap, and this makes the per-unit

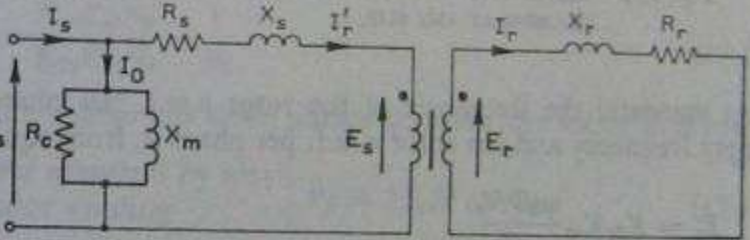


Fig. 13.1 EQUIVALENT CIRCUIT OF THE POLYPHASE INDUCTION MACHINE AT STANDSTILL

value of magnetizing current much higher than that of the power transformer. As a result the approximation of showing the shunt magnetizing branch of the equivalent circuit at the input terminals is less close than for the power transformer. The approximation is nevertheless acceptable for large machines, but not for small machines. To keep the magnetizing current as small as possible, the air-gap length of induction machines is made as short as is consistent with mechanical considerations.

A further difference between the polyphase induction machine and the power transformer is that in the former the windings are distributed, and this affects the effective turns ratio.

In this and subsequent sections it is assumed that the rotor has a 3-phase winding. A cage rotor is, in effect, a rotor with a large number of short-circuited phases. Such an arrangement may be represented by an equivalent 3-phase winding; I_r is not then the current in an actual rotor phase, but the stator current I_s is preserved as the true stator current.

The induced stator e.m.f. per phase when connected to a supply of frequency f hertz is, from eqn (11.20),

$$E_s = K_{ds}K_{ss}\frac{\omega_0 \Phi N_s}{\sqrt{2}} \tag{13.8}$$

where $\omega_0 = 2\pi f = 2\pi n_0 p$.

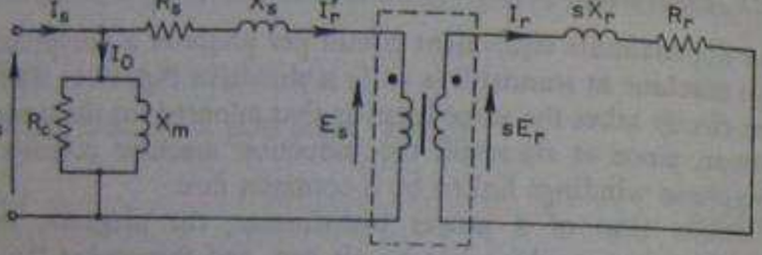


Fig. 13.2 EQUIVALENT CIRCUIT OF THE POLYPHASE INDUCTION MACHINE AT ANY SLIP, s

At standstill the frequency of the rotor e.m.f. per phase is the supply frequency and the rotor e.m.f. per phase is, from eqn (11.20),

$$E_r = K_{dr}K_{sr}\frac{\omega_0 \Phi N_r}{\sqrt{2}} \tag{13.9}$$

When the rotor rotates the rotor e.m.f. per phase is altered both in size and frequency.

$$\left.\begin{array}{l}\text{Rotor e.m.f. per phase}\\ \text{at any slip } s\end{array}\right\} = K_{dr}K_{sr}\frac{\omega_r \Phi N_r}{\sqrt{2}}$$

$$= K_{dr}K_{sr}\frac{s\omega_0 \Phi N_r}{\sqrt{2}}$$

$$= sE_r$$

The rotor reactance per phase at standstill is X_r. At any slip s, therefore, the rotor reactance per phase will be sX_r, since reactance is proportional to frequency. The equivalent circuit per phase of a polyphase inductor motor at any slip s is shown in Fig. 13.2.

13.3 Slip Ratios

The element enclosed by the dotted box in Fig. 13.2 represents an ideal or lossless induction machine. This differs from the ideal transformer considered in Chapter 9 in that (*a*) the current and voltage transformation ratios differ, and (*b*) the frequencies of the voltages and currents at the input and output terminal pairs of the ideal element also differ.

From eqns. (13.8) and (13.9) the effective turns ratio, k_t, at standstill ($s = 1$) is

$$k_t = \frac{E_s}{E_r} = \frac{K_{ds}K_{ss}N_s}{K_{dr}K_{sr}N_r} \tag{13.10a}$$

At any slip s the voltage ratio is

$$\frac{E_s}{sE_r} = \frac{k_t}{s} \tag{13.10b}$$

At any slip s, m.m.f., balance must exist between the stator and rotor phase windings so that

$$I_r' K_{ds}K_{ss}N_s = I_r K_{dr}K_{sr} N_r$$

or

$$\frac{I_r'}{I_r} = \frac{K_{dr}K_{sr}N_r}{K_{ds}K_{ss}N_s} = \frac{1}{k_t} \tag{13.11}$$

Assuming there are three phases on both the stator and rotor,

$$\left.\begin{array}{c}\text{Power absorbed by ideal}\\ \text{stator winding}\end{array}\right\} P_0 = 3E_sI_r' \cos\phi_r \tag{13.12}$$

This power is obtained from the supply when the machine acts as a motor, and from the prime mover driving the rotor when it acts as a generator.

$$\left.\begin{array}{c}\text{Power dissipated in}\\ \text{the rotor circuit}\end{array}\right\} P_r = 3sE_rI_r \cos\phi_r$$

$$= 3s\frac{E_s}{k_t}k_tI_r' \cos\phi_r$$

$$= 3sE_sI_r' \cos\phi_r \tag{13.13}$$

Dividing eqn. (13.12) by eqn. (13.13),

$$\frac{P_0}{P_r} = \frac{1}{s} \tag{13.14}$$

The power dissipated in the rotor circuit consists of winding loss in the rotor circuit and core loss in the rotor magnetic circuit. Since the core loss varies with frequency this implies that the equivalent circuit-element, R_r, is frequency dependent. Under normal running conditions, for plain induction motors, however, the rotor frequency and rotor core loss are low and the latter may usually be neglected. The power dissipated in the rotor is obtained from the ideal stator winding when the machine acts as a motor and from the prime mover when the machine acts as a generator.

When the machine acts as a motor the power absorbed by the ideal stator windings is greater than that dissipated in the rotor circuit except when the rotor is stationary (standstill) when they are equal. The difference in these two powers appears as gross mechanical power output:

$$\text{Mechanical power, } P_m = P_0 - P_r = P_0 - sP_0$$

or

$$P_m = P_0(1 - s) \tag{13.15}$$

Combining eqns. (13.14) and (13.15),

$$P_0 : P_r : P_m = 1 : s : (1 - s) \tag{13.16}$$

When the machine acts as a generator the net mechanical power input is the sum of the stator and rotor powers:

$$\text{Mechanical power, } P_m = P_0 + P_r$$

or

$$P_m = P_0(1 + s) \tag{13.17}$$

For generator action, therefore,

$$P_0 : P_r : P_m = 1 : s : (1 + s) \tag{13.18}$$

Eqn. (13.16) will serve for both motor and generator action if the slip s, the power absorbed by the ideal stator winding P_0 and the mechanical power P_m are taken to be negative for generator action, and it is remembered that P_m is the gross mechanical power output for the motoring mode and the net power input for the generating mode. Figs. 13.3(*a*) and (*b*) are block diagrams representing the power transfer in a plain induction machine for motor and generator action.

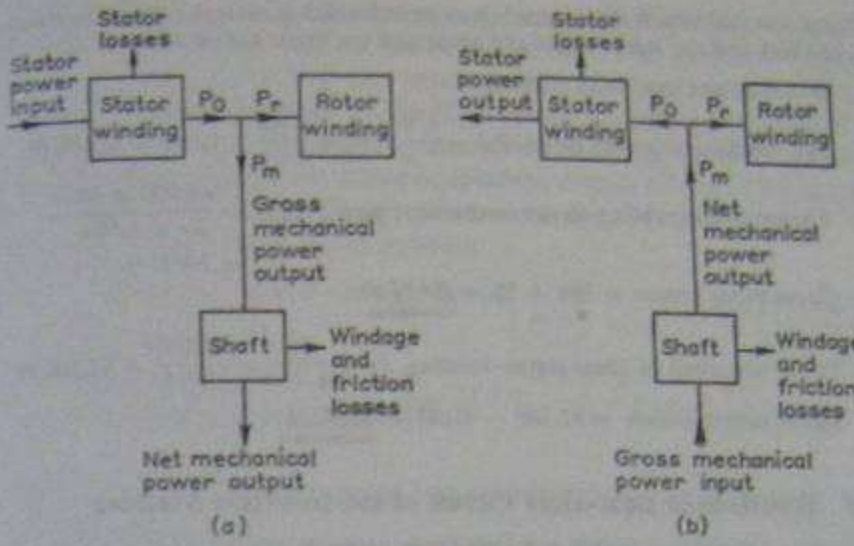


Fig. 13.3 POWER TRANSFER IN A PLAIN INDUCTION MACHINE
(*a*) Motoring mode (*b*) Generating mode

EXAMPLE 13.1 A 37·3 kW 4-pole 50 Hz induction machine has a friction and windage torque of 22 N-m. The stator losses equal the rotor circuit loss. Calculate:

(*a*) The input power to the stator when delivering full-load output at a speed of 1,440 rev/min.

(*b*) The gross input torque and stator output power when running at a speed of 1,560 rev/min. The stator losses are as in (*a*) and the windage and friction torque is unchanged.

(*a*) Synchronous speed $= \dfrac{f}{p} \times 60 = \dfrac{50 \times 60}{2} = 1{,}500$ rev/min

Per-unit slip, $s = \dfrac{n_0 - n_r}{n_r} = \dfrac{1{,}500 - 1{,}440}{1{,}500} = 0{\cdot}04$

Windage and friction loss $= 2\pi n_r T = \dfrac{2\pi \times 1{,}440 \times 22}{60} = 3{,}320$ W

Gross mechanical power output, $P_m = 37{,}300 + 3{,}320 = 40{,}620$ W

Power absorbed by ideal stator winding, $P_0 = \dfrac{P_m}{1 - s} = \dfrac{40{,}620}{0{\cdot}96} = 42{,}300$ W

Stator losses = Rotor loss $= sP_0 = 0{\cdot}04 \times 42{,}300 = 1{,}690$ W

Stator input power $= 42{,}300 + 1{,}690 = \underline{44{,}000 \text{ W}}$

(*b*) Per-unit slip $= \dfrac{1{,}500 - 1{,}560}{1{,}500} = -0{\cdot}04$

That is, the machine is now operating as an induction generator. Since the rotor circuit loss and the stator losses are equal and the latter are unchanged,

Rotor circuit loss, $P_r = 1{,}600\,\text{W}$

Net mechanical power input, $P_m = \dfrac{1+s}{s} P_r = \dfrac{1{\cdot}04}{0{\cdot}04} \times 1{,}690 = 44{,}000\,\text{W}$

Torque corresponding to net mechanical power input $= \dfrac{44{,}000 \times 60}{2\pi \times 1{,}560}$

$= 269\,\text{N-m}$

Gross input torque $= 269 + 22 = \underline{\underline{291\,\text{N-m}}}$

Power absorbed by ideal stator winding, $P_0 = \dfrac{P_m}{1+s} = \dfrac{44{,}000}{1{\cdot}04} = 42{,}300\,\text{W}$

Stator output power $= 42{,}300 - 1{,}690 = \underline{\underline{40{,}300\,\text{W}}}$

13.4 Transformer Equivalent Circuit of the Induction Machine

Referring to the equivalent circuit of Fig. 13.2, the rotor current per equivalent phase is

$$I_r = \frac{sE_r}{R_r + jsX_r}$$

If the numerator and denominator are divided by s this gives

$$I_r = \frac{E_r}{\dfrac{R_r}{s} + jX_r} \tag{13.19}$$

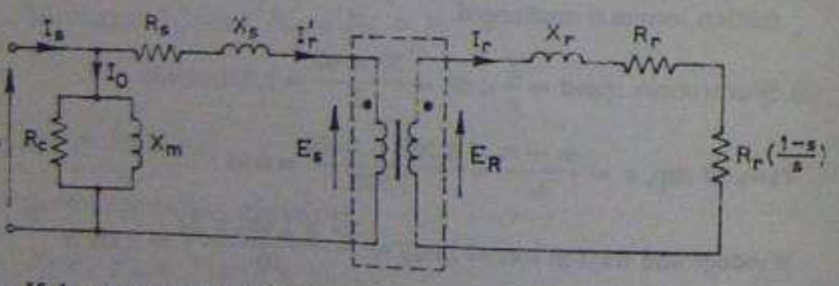


Fig. 13.4 TRANSFORMER EQUIVALENT CIRCUIT OF THE INDUCTION MACHINE

This latter expression for the rotor current per equivalent phase is consistent with the rotor equivalent circuit shown in Fig. 13.4. Although the value of I_r is unchanged this equivalent circuit is significantly different from that of Fig. 13.2 in that both the induced voltage and the reactance per equivalent rotor phase have their standstill ($s = 1$) values. Nor do the voltage and current ratios now differ as they did in the equivalent circuit of Fig. 13.1. Both are now equal to the virtual turns ratio, k_t. The element enclosed in the dotted box of Fig. 13.4 represents an ideal transformer. Therefore the power absorbed by the ideal stator winding must be equalled by the power delivered by the ideal rotor winding to R_r/s. Thus the power dissipated in the rotor equivalent circuit of Fig. 13.4 must be both the rotor loss and the gross mechanical power output. This may be shown to be so as follows.

$$\left.\begin{array}{l}\text{Added resistance per equivalent}\\ \quad\text{rotor phase}\end{array}\right\} = \frac{R_r}{s} - R_r = R_r\left(\frac{1-s}{s}\right)$$

$$\left.\begin{array}{l}\text{Power dissipated in}\\ \quad\text{added rotor resistance}\end{array}\right\} = 3I_r^2 R_r\left(\frac{1-s}{s}\right) = P_r\left(\frac{1-s}{s}\right) = P_m$$

That is, the equivalent circuit has additional resistance of $R_r\left(\frac{1-s}{s}\right)$ in each phase and the power dissipated in these additional resistances is equal to the gross mechanical power output. Fig. 13.4 shows the equivalent circuit of an induction machine which is also the equivalent circuit of a transformer the power dissipated in whose secondary load is equal to the gross mechanical power output of the induction machine when operating in the motoring mode.

Since the equivalent circuit of Fig. 13.4 is that of a transformer, the secondary values may be referred to the primary by multiplying them by k_t^2, the square of the virtual turns ratio. The turns ratio is defined in eqn. (13.10*a*). Fig 13.5 shows the referred equivalent circuit in which

$$\frac{R_r'}{s} = k_t^2\frac{R_r}{s} \quad \text{and} \quad X_r' = k_t^2 X_r$$

It should be noted that the equivalent circuits are valid only if the variations in speed or slip are relatively slow. Usually the moment of inertia of the rotor is sufficiently large for this condition to be realized.

13.5 Torque developed by an Induction Machine

An expression for the torque developed by an induction machine may be obtained by reference to the equivalent circuit of Fig. 13.5. Assuming that the stator winding has three phases,

$$\text{Rotor circuit loss, } P_r = 3(I_r')^2 R_r' \tag{13.20}$$

$$\left.\begin{array}{l}\text{Power absorbed by ideal}\\ \quad\text{stator winding}\end{array}\right\} P_0 = \frac{P_r}{s} = 3(I_r')^2\frac{R_r'}{s} \tag{13.21}$$

$$\left.\begin{array}{l}\text{Gross mechanical}\\ \text{power output}\end{array}\right\} P_m = 3(I_r')^2 R_r' \frac{1-s}{s} \tag{13.22}$$

$$\text{Gross torque developed, } T = \frac{3}{2\pi n_r}(I_r')^2 R_r' \frac{1-s}{s} \tag{13.23}$$

where $n_r = (1-s)n_0$ (13.5)

Substituting for n_r in eqn. (13.23),

$$T = \frac{3}{2\pi n_0}(I_r')^2 \frac{R_r'}{s} \tag{13.24}$$

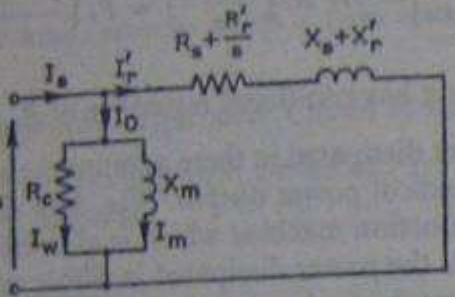


Fig. 13.5 REFERRED TRANSFORMER EQUIVALENT CIRCUIT OF THE INDUCTION MACHINE

Comparing eqn. (13.23) with eqn. (13.21) it will be seen that

$$T = \frac{P_0}{2\pi n_0} \tag{13.25}$$

Thus the torque developed is proportional to the power, P_0, absorbed by the ideal stator winding . The quantity P_0 is sometimes referred to as the torque measured in "synchronous watts", which presumably implies that, if this power is divided by the synchronous angular velocity, the torque is obtained.

Referring to the equivalent circuit of Fig. 13.5, evidently

$$I_r' = \frac{V_s}{\sqrt{\left\{\left(R_s + \frac{R_r'}{s}\right)^2 + (X_s + X_r')^2\right\}}}$$

Substituting for $(I_r')^2$ in eqn. (13.24),

$$T = \frac{3}{2\pi n_0} \frac{V_s^2}{\left(R_s + \frac{R_r'}{s}\right)^2 + (X_s + X_r')^2} \frac{R_r'}{s} \tag{13.26}$$

Multiplying numerator and denominator by s gives

$$T = \frac{3V_s^2}{2\pi n_0} \frac{sR_r'}{(sR_s + R_r')^2 + s^2(X_s + X_r')^2} \tag{13.27}$$

If the stator impedance is neglected, this equation reduces to

$$T = \frac{3V_s^2}{2\pi n_0} \frac{sR_r'}{(R_r')^2 + s^2(X_r')^2} \tag{13.28}$$

Eqns. (13.27) and (13.28) have been obtained by considering motor action, but they apply equally to generator action. If the torque is taken to be positive when the machine acts as a motor it will be negative for generator action since the slip then becomes negative. For motor action eqn. (13.28) gives the gross torque developed, but for generator action it gives the net input torque, as will be evident from a consideration of Figs. 13.2(*a*) and (*b*), the torque in each case being $P_m/2\pi n_r$.

13.6 Slip/Torque Characteristics of the Induction Machine

Referring to the expression for torque given by eqn. (13.27), since the slip s is positive in both the numerator and the denominator, the

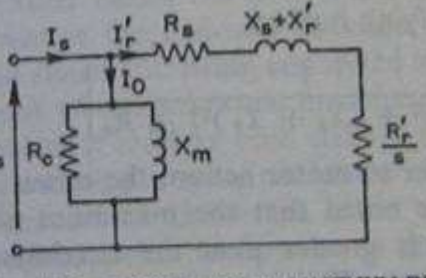


Fig. 13.6 PERTAINING TO MAXIMUM DEVELOPED TORQUE

torque will be zero both at $s = 0$ and $s = \infty$. Therefore, the torque will be a maximum at some intermediate value of slip.

From eqn. (13.24),

$$T = \frac{3}{2\pi n_0} (I_r')^2 \frac{R_r'}{s} \tag{13.24}$$

Therefore the torque will be a maximum when there is maximum power transfer into the load R_r'/s. As shown in Fig. 13.6, the source impedance is $R_s + j(X_s + X_r')$ assuming the supply itself to have zero internal impedance. According to Section 2.4 the maximum power is transferred to R_r'/s when

$$\frac{R_r'}{s} = \pm\sqrt{\{R_s^2 + (X_s + X_r')^2\}}$$

or

$$s = \pm\frac{R_r'}{\sqrt{\{R_s^2 + (X_s + X_r')^2\}}} \tag{13.29}$$

The plus sign refers to motor action, the minus sign to generator action.

An expression for maximum torque may be obtained by substituting in eqn. (13.27) the value of s given in eqn. (13.29). The algebra is a little simplified if in the first instance a substitution for $s^2(X_s + X_r')^2$ is made. From eqn. (13.29),

$$s^2(X_s + X_r')^2 = (R_r')^2 - s^2R_s^2$$

In eqn. (13.27),

$$T_{max} = \frac{3V_s^2}{2\pi n_0}\frac{sR_r'}{(sR_s + R_r')^2 + (R_r')^2 - s^2R_s^2}$$

$$= \frac{3V_s^2}{2\pi n_0}\frac{1}{2}\frac{s}{(sR_s + R_r')}$$

Substituting for s and simplifying further,

$$T_{max} = \pm\frac{3V_s^2}{2\pi n_0}\frac{1}{2}\left[\frac{1}{\sqrt{\{R_s^2 + (X_s + X_r')^2\}} \pm R_s}\right] \tag{13.30}$$

Here again the plus signs refer to motor action, the minus signs to generator action. It is to be noted that the maximum net input torque for generator action is greater than the maximum gross output torque for motor action. This inequality would disappear if the stator resistance were negligible. Although, as eqn. (13.29) shows, the slip at which the maximum torque occurs is proportional to the referred value of rotor resistance per stator phase, R_r', the actual value of the maximum torque in independent of R_r'. Therefore variation of R_r' changes the slip at which maximum torque occurs without affecting its value. The maximum torque is sometimes called the *pull-out torque*.

If stator impedance is neglected (i.e. $R_s = 0$, $X_s = 0$) eqn. (13.29) becomes

$$s = \pm\frac{R_r'}{X_r'} \tag{13.31}$$

and eqn. (13.30) becomes

$$T_{max} = \pm \frac{3}{2\pi n_0}\frac{1}{2X_r'} \tag{13.32}$$

It will be appreciated that, to obtain a high starting torque and a high maximum torque, the combined rotor and stator leakage reactance must be small. The shorter the air-gap is made the more the leakage flux is reduced. This is an additional reason for minimizing the air-gap.

Fig. 13.7 is a typical slip/torque characteristic of an induction machine. The hatched areas in the region of $s = 0$ show the normal operating range of the machine for motor and generator action. Referring to motor action, AB represents the torque at standstill or starting torque. Provided the load torque is less than this the motor will accelerate until the developed motor torque and load torque come into equality at a speed close to but less than synchronous speed. The machine operates stably as a motor over the range indicated. If the machine is operating in this region and a load torque in excess of the maximum motoring torque, CD, is imposed on the machine, it will decelerate to standstill or stall. The range DB represents unstable motor action.

Fig. 13.7 shows positive values of slip greater than unity. To achieve such values the rotor must be coupled to a prime mover and driven in the opposite direction to that of the stator rotating field, the stator still being connected to the 3-phase supply. In such conditions of operation the machine acts as neither a motor nor a generator as it receives both electrical and mechanical input power, all the power input being dissipated as loss. This mode of operation is referred to as *brake action*.

For the machine to operate as an induction generator a prime mover must drive the rotor in the same direction as the stator rotating field but at a higher speed with the stator connected to a pre-existing supply. In Fig. 13.7 OF represents the range of stable generator action. If the input torque to the generator exceeds the maximum generating torque, EF, the machine accelerates and passes into the region of unstable generator action, FG. Unless the input torque is removed, the speed may rise dangerously.

At some negative value of slip the machine will pass from unstable generator action to brake action. This will occur when the stator I^2R loss exceeds the power absorbed by the ideal stator winding, since the machine must then draw power from the electrical supply to meet completely the stator loss as well as having mechanical power input.

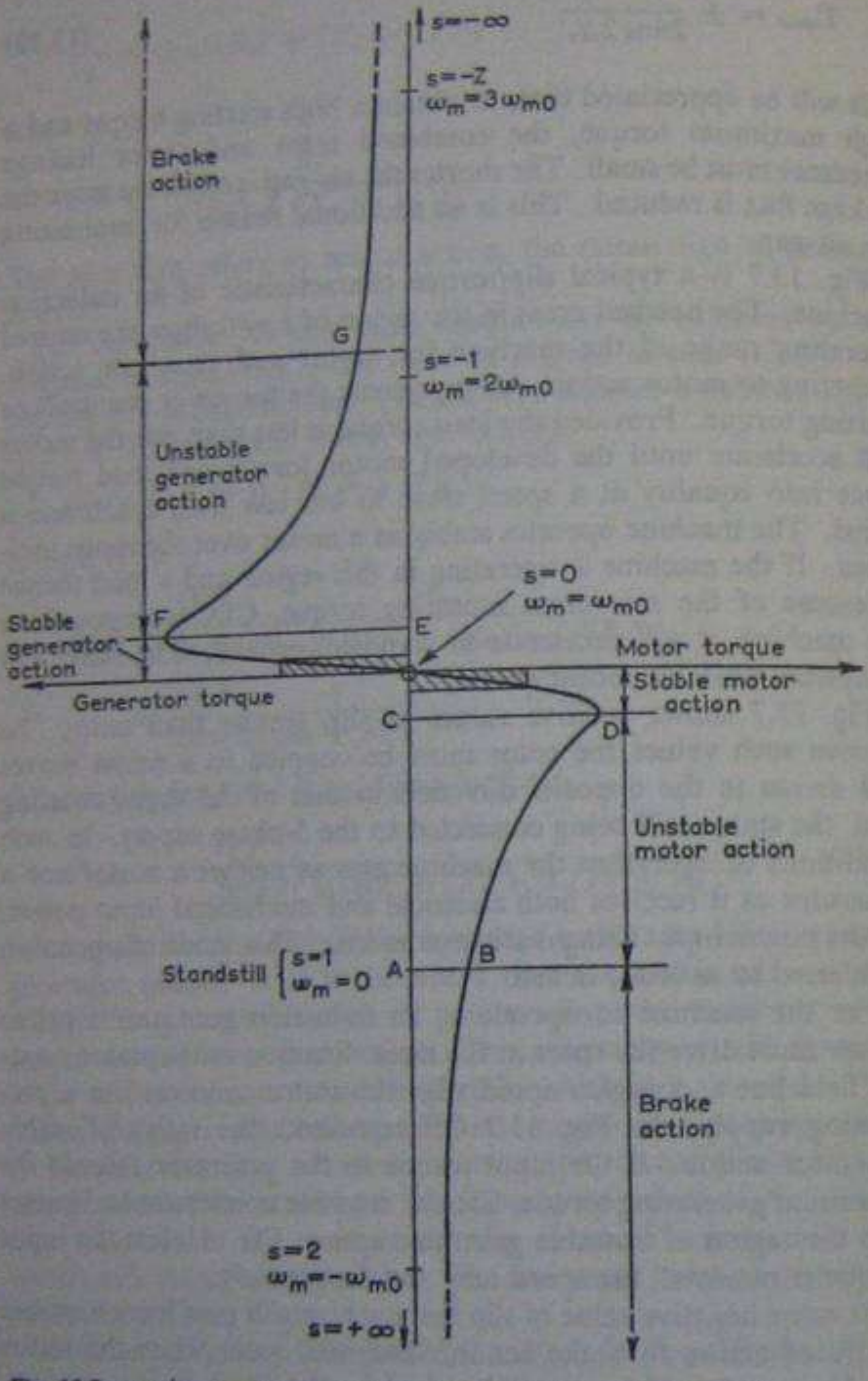


Fig. 13.7 SLIP/TORQUE CHARACTERISTIC OF THE INDUCTION MACHINE

$$\text{Stator } I^2R \text{ loss} = 3(I_r')^2R_s \tag{13.33}$$

$$\text{Power absorbed by ideal stator, } P_0 = -3(I_r')^2\frac{R_r'}{s} \tag{13.21}$$

P_0 is taken to be negative when the input to the ideal stator winding is derived from a mechanical source. Equating eqns. (13.33) and (13.21),

$$s = -\frac{R_r'}{R_s} \tag{13.34}$$

At this value of slip the machine changes from unstable generator to brake action. Since R_r' and R_s will be of the same order this value of slip will be approximately -1.

The slip/torque characteristic shown in Fig. 13.7 has its maximum value at a relatively small value of slip. This is typical of induction machines and is desirable, since when the machine operates as a motor the speed regulation with load is small, giving the machine a speed/torque characteristic over its operating region similar to that of the d.c. shunt motor. This matter is discussed further in Section 13.7. Further, since from eqn. (13.14) the slip is equal to $s = P_r/P_0$, an induction machine operating with a large value of slip would have a large rotor loss and consequently a low efficiency. A disadvantage of having the maximum torque occur at a low value of slip is that, as Fig. 13.7 shows, this arrangement makes the starting torque low.

EXAMPLE 13.2 A 440 V 4-pole 3-phase 50 Hz slip-ring induction motor has its stator winding mesh connected and its rotor winding star connected. The standstill voltage measured between slip rings with the rotor open-circuited is 218 V. The stator resistance per phase is 0·6 Ω and the stator reactance per phase is 3 Ω. The rotor resistance per phase is 0·05 Ω and the rotor reactance per phase is 0·25 Ω. Calculate the maximum torque and the slip at which it occurs. If the ratio of full-load to maximum torque is 1 : 2·5 find the full-load slip and the power output.

All values must be referred to either the stator or the rotor. It is usual to refer to the stator. Since the rotor is star connected:

$$\text{Induced standstill rotor voltage, } E_r = \frac{218}{\sqrt{3}} = 126\,\text{V}$$

From eqn. (13.10*a*) the standstill turns ratio is

$$k_t = \frac{E_s}{E_r} = \frac{440}{126} = 3{\cdot}49$$

Rotor resistance/phase referred to stator, $R_r' = 0{\cdot}05 \times 3{\cdot}49^2 = 0{\cdot}61\,\Omega$

Rotor reactance/phase referred to stator, $X_r' = 0{\cdot}25 \times 3{\cdot}49^2 = 3{\cdot}05\,\Omega$

From eqn. (13.29) the slip for maximum torque is

$$s = \frac{R_r'}{\sqrt{((R_s')^2 + (X_s \times X_r')^2)}} = \frac{0{\cdot}61}{\sqrt{(0{\cdot}6^2 + 6{\cdot}05^2)}} = \underline{\underline{0{\cdot}1}}$$

$$\text{Synchronous speed, } n_0 = \frac{f}{p} = \frac{50}{2} = 25\,\text{rev/s}$$

From eqn. (13.30) the maximum torque is

$$T_{max} = \frac{3V_s^2}{2\pi n_0}\frac{1}{2}\left[\frac{1}{\sqrt{\{R_s^2 + (X_s + X_r')^2\}} + R_s}\right]$$

$$= \frac{3 \times 440^2}{2\pi 25}\frac{1}{2}\frac{1}{6{\cdot}06 + 0{\cdot}6} = \underline{\underline{278\,\text{N-m}}}$$

$$\text{Full-load torque} = \frac{278}{2{\cdot}5} = 111\,\text{N-m}$$

The slip for full-load torque may be obtained from eqn. (13.27):

$$T = \frac{3V_s^2}{2\pi n_0}\frac{sR_r'}{(sR_s + R_r')^2 + s^2(X_s + X_r')^2} \tag{13.27}$$

i.e.

$$111 = \frac{3 \times 440^2}{2\pi \times 25}\frac{0{\cdot}61s}{(0{\cdot}6s + 0{\cdot}61)^2 + 6{\cdot}05^2s^2}$$

This gives

$$s^2 - 0{\cdot}53s + 0{\cdot}01 = 0$$

so that

$$s = \frac{0{\cdot}53 \pm \sqrt{(0{\cdot}53^2 - 4 \times 0{\cdot}01)}}{2} = \underline{\underline{0{\cdot}02 \text{ or } 0{\cdot}51}}$$

Since the slip for maximum torque is 0·1, $s = 0{\cdot}02$ is on the stable part of the slip/torque characteristic and $s = 0{\cdot}51$ is on the unstable part. Selecting the value of s giving stable operation,

$$\text{Power output, } P_m = 2\pi n_r T = 2\pi \times 25(1 - 0{\cdot}02) \times 111 = \underline{\underline{17{\cdot}1\text{ kW}}}$$

13.7 Starting

SLIP-RING MACHINES

To obtain a satisfactory operating characteristic giving a reasonable efficiency and a small speed regulation with load, the slip for the maximum torque developed by an induction motor must have a value in the range from 0·1 to 0·2, as explained at the end of Section 13.6. From eqn. (13.29) the slip for maximum torque is

$$s = \frac{R_r'}{\sqrt{\{R_s^2 + (X_s + X'_r)^2\}}} = \text{from } 0{\cdot}1 \text{ to } 0{\cdot}2$$

At starting $s = 1$, and to obtain maximum torque on starting,

$$\frac{R_r'}{\sqrt{\{R_s^2 + (X_s + X'_r)^2\}}} = 1$$

In the plain-cage-rotor induction motor these conflicting requirements cannot well be met, though specially shaped rotor slots or a double-cage rotor may make this possible. In slip-ring machines,

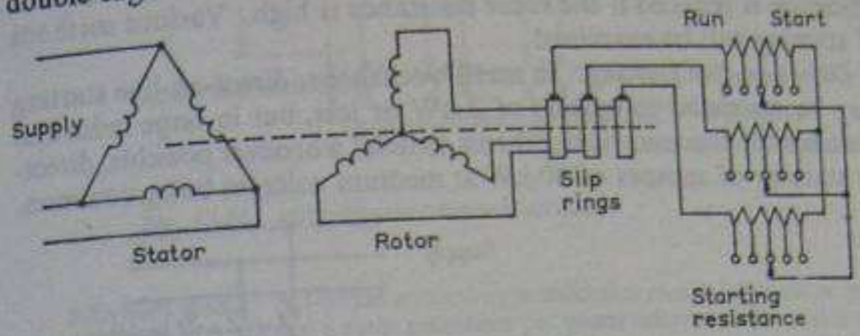


Fig. 13.8 STARTING OF WOUND-ROTOR MACHINES

however, the rotor resistance per phase is such as to give a satisfactory operating characteristic.

Slip-ring machines are invariably started by means of external resistances connected through the slip rings to the rotor circuit (Fig. 13.8). The machine is started with all the resistances in, giving a high starting torque. As the machine runs up to speed the

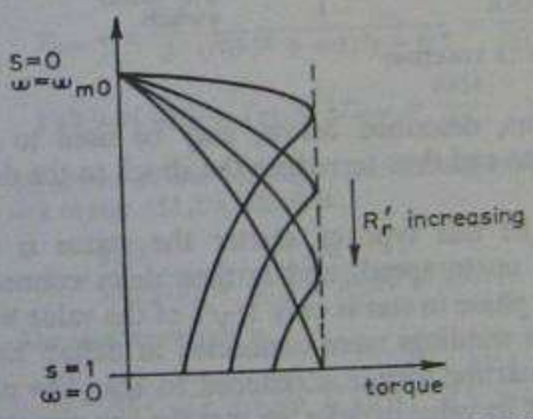


Fig. 13.9 SLIP/TORQUE CURVES FOR ROTOR-RESISTANCE STARTING

external resistance is reduced until the machine attains full speed with no external resistance.

Fig. 13.9 shows the slip/torque curves of a slip-ring induction motor corresponding to various positions of the starting resistance.

NON-SLIP-RING MACHINES

Stator starting must be used for cage-rotor machines, since no connexion can be made to the rotor, and direct switching of large machines would cause huge starting currents, which must be avoided. The cage-rotor machine suffers from the disadvantage that the starting torque is low if the resistance is low, while the efficiency is reduced if the rotor resistance is high. Various methods of starting will be examined.

Direct-on-line starting. In small workshops, direct-on-line starting may be restricted to motors of 2 kW or less, but in large industrial premises the tendency is to "direct switch" whenever possible, direct-on starting of motors of 40 kW at medium voltages being common.

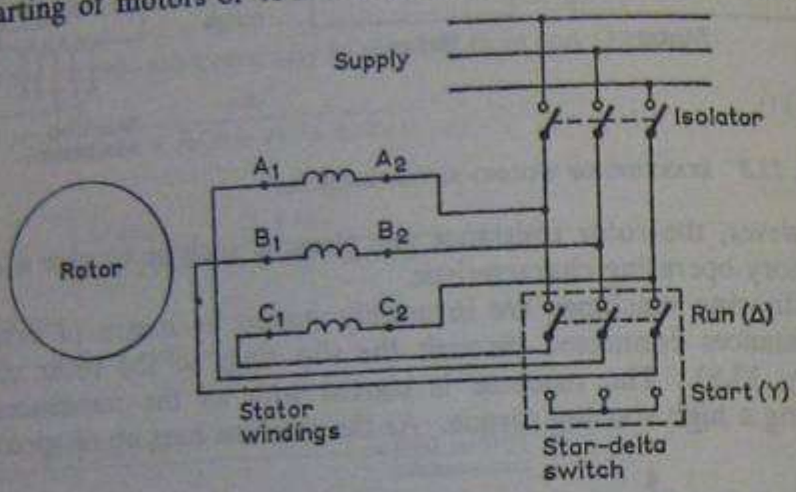


Fig. 13.10 STAR-DELTA STARTING

Reduced-voltage starters, described below, may be used to limit the initial starting torque and thus to reduce the shock to the driven machine.

Star-delta starting. In this type of starter the stator is star-connected for running up to speed, and is then delta connected. The applied voltage per phase in star is only $1/\sqrt{3}$ of the value which would be applied if the windings were connected in delta; hence, from eqn. (13.27), the starting torque is reduced to 1/3. The phase current in star is $1/\sqrt{3}$ of its value in delta, so that the line current for star connexion is 1/3 of the value for delta. Fig. 13.10 shows a connexion diagram for a star-delta starter.

Auto-transformer starting. In auto-transformer starting the transformer has at least three tappings giving open-circuit voltages of not less than 40, 60 and 75 per cent of line voltage for starting, and the stator is switched directly to the mains when the motor has run up to speed (Fig. 13.11). If the fractional tapping is x, then the applied voltage per phase on starting is xV_1 (where V_1 is the mains voltage), and the starting torque is reduced by x^2. The starting current from the mains will also be reduced by approximately x^2.

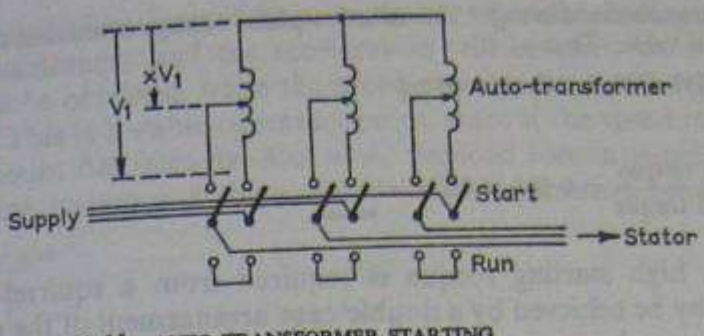


Fig. 13.11 AUTO-TRANSFORMER STARTING

EXAMPLE 13.3 A 3-phase squirrel-cage induction motor has a stator resistance per phase of 0·5Ω and a rotor resistance per phase referred to the stator of 0·5Ω. The total standstill reactance per phase referred to the stator is 4·92Ω. If the ratio of maximum torque to full-load torque is 2:1, find the ratio of actual starting to full-load torque for (*a*) direct starting, (*b*) star-delta starting and (*c*) auto-transformer starting with a tapping of 75 per cent.

The maximum torque is given by eqn. (13.30):

$$T_{max} = \pm \frac{3V_s^2}{2\pi n_0}\frac{1}{2}\left[\frac{1}{\sqrt{\{R_s^2 + (X_s + X_r')^2\}} \pm R_s}\right]$$

For motor action

$$T_{max} = k\frac{V_s^2}{2}\frac{1}{\sqrt{(0{\cdot}5^2 + 4{\cdot}92^2)} + 0{\cdot}5} = \frac{kV_s^2}{10}$$

$$\text{Full-load torque, } T_{FL} = \tfrac{1}{2}T_{max} = \frac{kV_s^2}{20}$$

(*a*) *Direct-on-line starting*. The starting torque is obtained by substituting $s = 1$ in eqn. (13.27), which is

$$T = \frac{3V_s^2}{2\pi n_0}\frac{sR_r'}{(sR_s + R_r')^2 + s^2(X_s + X_r')^2} \tag{13.27}$$

The starting torque is

$$T_0 = kV_s^2\frac{0{\cdot}5}{(0{\cdot}5 + 0{\cdot}5)^2 + 4{\cdot}92^2} = \frac{kV_s^2}{2 \times 24{\cdot}2}$$

Therefore

$$\frac{\text{Starting torque}}{\text{Full-load torque}} = \frac{T_0}{T_{FL}} = \frac{kV_s^2}{2 \times 24{\cdot}2} \times \frac{20}{kV_s^2} = \underline{\underline{0{\cdot}413}}$$

(*b*) *Star-delta starting*. The effective phase voltage is reduced to $1/\sqrt{3}$ of its original value. Therefore

$$T_0 = \left(\frac{1}{\sqrt{3}}\right)^2 \text{ of } T_0 \text{ for direct starting, and}$$

$$\frac{\text{Starting torque}}{\text{Full-load torque}} = \frac{0{\cdot}393}{3} = \underline{\underline{0{\cdot}131}}$$

(c) *Auto-transformer starting*. The effective phase voltage is reduced to 0·75 of its original value. Thus

$T_0 = 0{\cdot}75^2$ of T_0 for direct starting

and

$$\frac{\text{Starting torque}}{\text{Full-load torque}} = 0{\cdot}393 \times 0{\cdot}75^2 = \underline{\underline{0{\cdot}221}}$$

Where a high starting torque is required from a squirrel-cage motor, it may be achieved by a double-cage arrangement of the rotor conductors, as shown in Fig. 13.12(*a*). The equivalent electrical

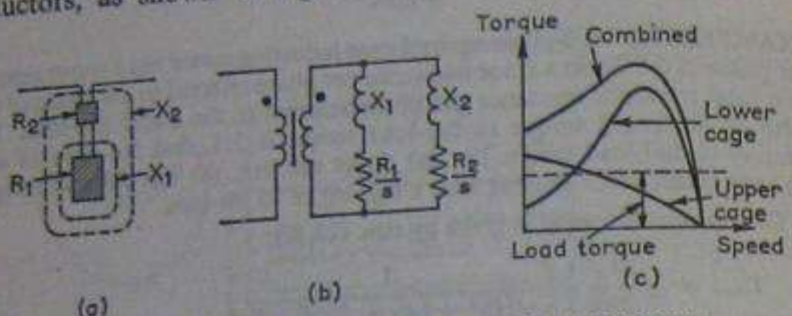


Fig. 13.12 IMPROVEMENT IN STARTING TORQUE OF CAGE ROTORS
(*a*) Double-cage rotor
(*b*) Equivalent circuit
(*c*) Combined torque/speed characteristics

rotor circuit is shown at (*b*), where X_1 and X_2 are leakage reactances. This equivalent circuit neglects mutual inductance between the cages. For the upper cage the resistance is made intentionally high, giving a high starting torque, while for the lower cage the resistance is low, and the leakage reactance is high, giving a low starting torque but high efficiency on load. The resultant characteristic will be roughly the sum of these two as shown in Fig. 13.12(*c*).

If a 3-phase induction motor starts in the wrong direction, this can be remedied by interchanging any two of the three supply leads to the stator.

13.8 Stability and Crawling

Curve (*a*) in Fig. 13.13 is the torque/speed curve for a typical induction motor. Consider that this motor is required to drive a constant-torque load having the torque/speed characteristic illustrated by curve (*b*). T_0, the starting torque with direct-on switching, is greater than the load torque T_b; thus there will be an excess starting torque $(T_0 - T_b)$ which will accelerate the motor and the load. The acceleration at any speed will be proportional to the torque difference $(T - T_b)$

so that the acceleration will be a maximum when the driving torque, T, is a maximum; and the acceleration will be zero, i.e. a steady speed will be obtained, when the speed corresponds to the operating point A. This is a stable operating point, since if the speed rose by a small amount Δn_r from its value at A, the load torque would exceed the driving torque and there would be a deceleration back to the

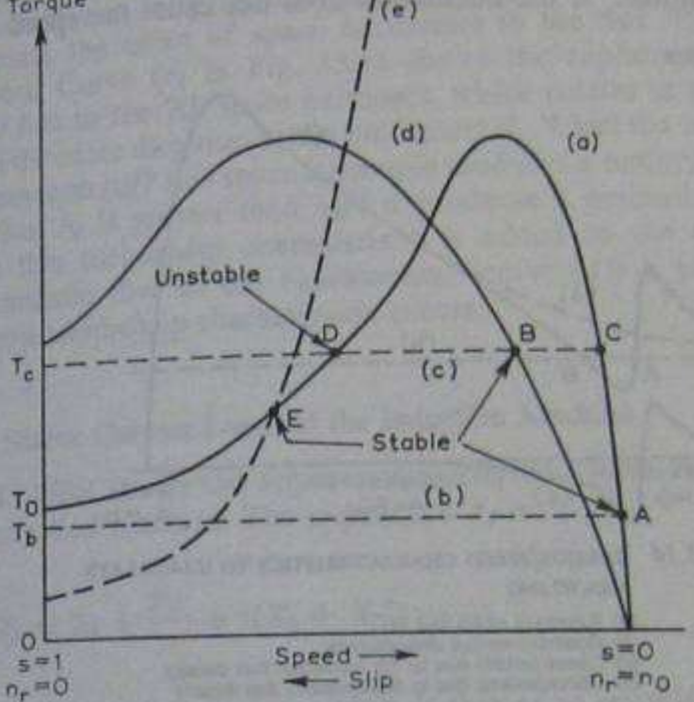


Fig. 13.13 PERTAINING TO STABILTY

--- Load characteristics
—— Motor characteristics

speed at A, and vice versa for a decrease in speed. The conclusions from this argument are:

(*a*) The operating point must be at the intersection of the two torque/speed characteristics.

(*b*) The slope of the load torque/speed curve must be greater than that of the driving-motor torque/speed curve for the operating point to be stable; i.e.

$$\frac{dT}{dn_r} \text{ for the load} > \frac{dT}{dn_r} \text{ for the drive}$$

Curve (*c*) in Fig. 13.13 represents a second load. In this case the load torque T_c is greater than the starting torque T_0 of the motor.

and with direct-on switching the motor would fail to start. The motor could be started by the use of additional rotor resistance sufficient to give the motor the characteristic of curve (*d*). The motor would drive the load at the speed corresponding to point C when the additional rotor resistance is short-circuited. Operation at point C may be unsatisfactory, since C is relatively near the maximum torque point and fluctuations in the load might too easily stall the motor. If the fluctuation does not cause the speed to fall

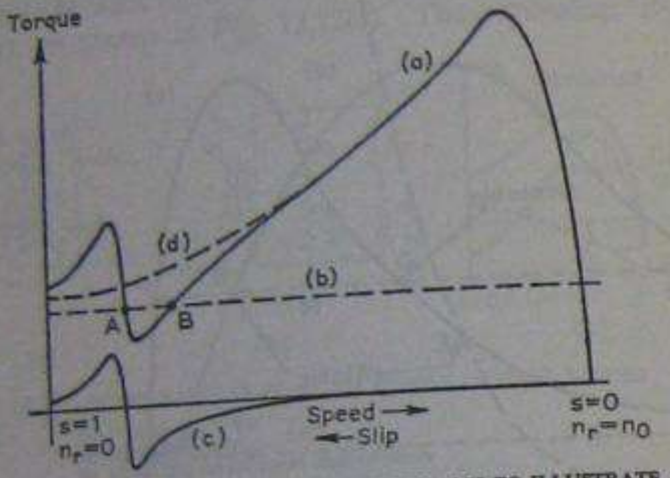


Fig. 13.14 TORQUE/SPEED CHARACTERISTICS TO ILLUSTRATE CRAWLING

(*a*) Resultant of (*c*) and (*d*)
(*b*) Constant-torque characteristic
(*c*) Characteristic due to 7th harmonic flux density
(*d*) Characteristic due to fundamental flux density

below that at D the motor should accelerate the load back to its speed at C when the load torque returns to normal. Since dT/dn_r for the load is not greater than dT/dn_r for the drive at D, this is an unstable operating point; i.e. if some random cause makes the speed fall slightly the load torque will exceed the driving torque and cause a further reduction in speed, or vice versa. The portion of the normal characteristic curve (*a*) which lies to the left of the maximum-torque point is called the *nominally unstable* portion. Though it is not normally possible for a motor to operate at a point on the nominally unstable portion of its characteristic, this may be arranged if a load, such as a fan, with a steep rising characteristic is chosen. A particular case is represented by curve (*e*); the motor would drive this load at a speed corresponding to the point E.

An induction motor may sometimes run in a stable manner at a low speed on a constant-torque load. This can be the result of a kink in the normal torque/speed characteristic. In Fig. 13.14

curve (*a*) shows a torque/speed curve with such a kink, and curve (*b*) represents a constant-torque load. The intersection A of curves (*a*) and (*b*) represents a stable operating point, so that the machine would not run up to full speed but merely drive the load at the speed corresponding to point A. This is termed *crawling*. To make the motor run up to full speed the load would have to be reduced to a value less than that of the minimum occurring between A and B. The kinks are due to irregularities (such as teeth) in the machine which accentuate the effect of space harmonics in the flux density distribution. Curve (*c*) in Fig. 13.14 shows the slip/torque characteristic due to the 7th space harmonic, which rotates at a speed of $n_0/7$ in the same direction as the fundamental. When the rotor speed n_r is less than $n_0/7$ this space harmonic produces a motoring torque, but when n_r is greater than $n_0/7$ it produces a generating torque. When this torque/slip characteristic is added to the torque/slip characteristic due to the fundamental (curve (*d*)) a kink in the resultant torque/slip characteristic occurs.

Learning Outcome 3.2

Contrast the performance of the various reduced voltage motor starting techniques

Learning Outcome 3.3

Draw the power and control circuitry for the various types of induction motor starters including open and closed transition

REDUCED-VOLTAGE STARTING METHODS

Reduced-voltage starters operate such that input current and consequently torque are reduced during starting. Table 12–2 briefly describes the various methods of starting and gives features and limitations of each.

When motors are started at reduced voltage, the current at the motor terminals is reduced in direct proportion to the voltage reduction, while the torque is reduced as the *square* of the voltage reduction. For example, if the "typical" motor were started at 65% of line voltage, the starting current would be 42% and the torque would be 42% of full-voltage values. Thus reduced-voltage starting provides an effective means of reducing both current and torque (Fig. 12–19).

PRIMARY RESISTOR STARTING

In primary resistor starting, a resistor is connected in each motor line (in one line only in single-phase starters) to produce a voltage drop due to the motor starting current. A timing relay shorts out the resistors

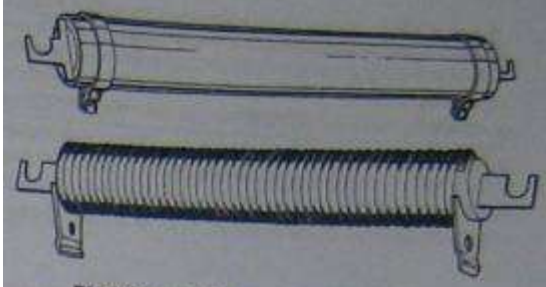


FIGURE 12-16 Wire-wound resistors used in primary resistor starter circuits. (Westinghouse)

after the motor has accelerated. Thus the motor is started at reduced voltage but operates at line voltage.

Figure 12-16 shows two types of motor starter resistors. The resistance element will retain its mechanical and electrical properties both during and after repeated heating and cooling. All metal parts are either plated with or fabricated of corrosion-resistant material for overall corrosion protection. Under certain conditions operating temperatures may reach 600°C and not change the resistance value. These are 11, 14, 17, and 20 in. long and come in wattage ratings of 450 to 1320. Table 12-3 shows the resistance ranges and other factors. Note the current-handling ability of the resistors.

Primary resistor starters are sometimes known as *cushion* starters. The main reason for the name is their ability to produce a smooth, cushioned acceleration with closed transition. However, this method is not as efficient as other methods of reduced-voltage starting, but it is ideally suited for applications such as conveyors, textile machines, or other delicate machinery where reduction of starting torque is of prime consideration.

Operation

Figure 12-17 is the reduced-voltage magnetic starter that uses resistors to operate a three-phase motor properly at start. Closing the START button or other pilot device energizes the start contactor (S) shown in Fig. 12-18. This connects the motor in series with the starting resistors for a reduced-voltage start. The contactor (S) is now sealed in through its interlock (Sa). Timing relay (TR) is energized, and after a preset time interval its contacts (TR_{TC}) close. This energizes the run contactor, RUN, which seals through its interlock (RUN_a). The contacts (RUN) close, bypassing the starting resistors, and the motor will now running at full voltage. The contactor (S) and timing relay (TR) are deenergized when the interlock (RUN_b) opens.

An overload, which opens the STOP button or other pilot device, deenergizes the (RUN) contactor. This removes the motor from the line.

TABLE 12-2 STARTING METHOD CHARACTERISTICS*

Starting Method	Operation	Tap	Starting Current (% of locked rotor current)	Starting Torque (% of locked rotor torque)	Open or Closed Transition	Basic Characteristics: Advantages	Basic Characteristics: Limitations
Across-the-line	Initially connects motor directly to power lines.		100%	100%	None	1. Lowest cost. 2. Highest starting torque. 3. Used with any standard motor. 4. Least maintenance.	1. High starting current. 2. High starting torque. 3. May shock driven machine.
Primary resistance reduced voltage	Inserts resistance units in series with motor during first step(s).		50–80%	25–64%	Closed	1. Smoothest starting. 2. Least shock to driven machine. 3. Most flexible in application. 4. Used with any standard motor.	1. High power loss because of heating resistors. 2. Heat must be dissipated. 3. Low torque per ampere input. 4. Highest cost.
Autotransformer reduced voltage	Uses autotransformer to reduce voltage applied to motor.	50% 65% 80%	25% 42% 64%	25% 42% 64%	Closed	1. Best for hard to start loads. 2. Adjustable starting torque. 3. Used with any standard motor. 4. Less strain on motor.	1. May shock driven machine. 2. High cost.
Wye–Delta	Starts motor with windings wye connected, then reconnects them in delta connection for running.		33%	33%	Open or closed	1. Medium cost. 2. Low starting current. 3. Low starting torque. 4. Less strain on motor.	1. Low starting torque. 2. Requires delta-wound motor.
Part Winding	Starts motor with only part of windings connected, then adds remainder for running.		70–80%	50–60% Minimum pull-up torque 35% of full-load torque.	Closed	1. Low cost. 2. Popular method for medium starting torque applications. 3. Low maintenance.	1. Not good for frequent [illegible] 2. May require special wound motor. 3. Low pull-up torque. 4. May not come up to speed on first step when started with load applied.

NOTE: The reduced starting torque (LRT) indicated in this table for the various reduced starting methods can prevent starting [illegible] inertia loads and must be considered when sizing motors and choosing starters.

TABLE 12-3 RESISTOR RANGES AND PROPERTIES

	Low R-High Current			High R-Low Current		
Unit Length (in.)	Resistance Range (Ω)	Current Range (A)	Heat Dissipation (Watts per unit*)	Resistance Range (Ω)	Current Range (A)	Heat Dissipation (Watts per unit*)
11	0.051–4.3	11–104	450–630	4.0–2000	0.46–10.3	426
14	0.069–5.7	11–104	620–820	5.0–2500	0.48–10.8	575
17	0.085–7.1	11–104	770–1080	5.0–2500	0.53–12.0	700
20	0.10–8.6	11–104	900–1320	6.4–4000	0.47–11.8	900

*Approximate only.

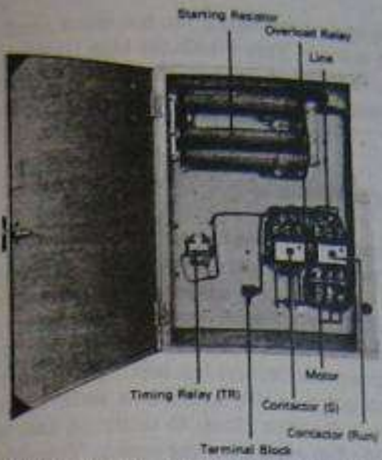


FIGURE 12–17 *Primary resistor type of magnetic starter. (Westinghouse)*

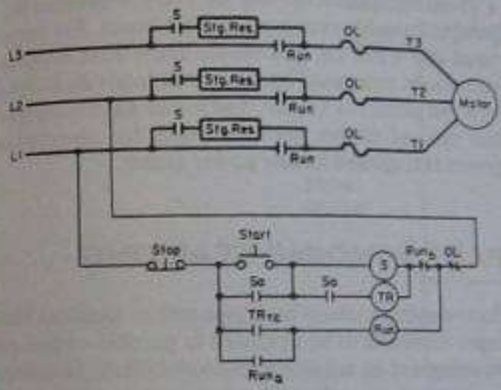


FIGURE 12–18 *Wiring diagram for a primary resistor type of starter. (Westinghouse)*

Primary resistor starters provide extremely smooth starting due to the increasing voltage across the motor terminals as the motor accelerates. Since motor current decreases with increasing speed, the voltage drop across the resistor decreases as the motor accelerates—and the motor terminal voltage increases. Thus if a resistor is shorted out as the motor reaches maximum speed, there is little or no increase in current or torque.

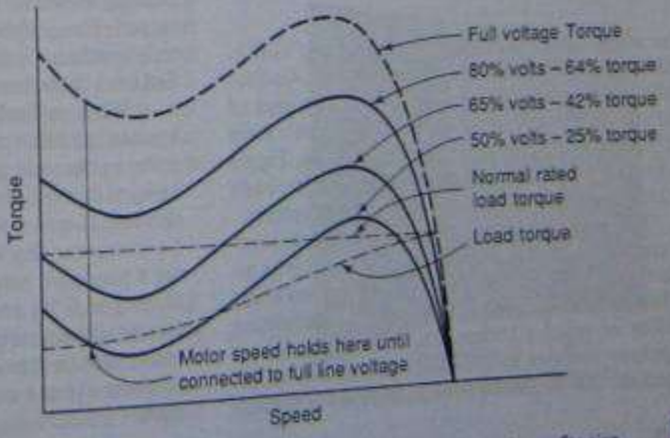
AUTOTRANSFORMER STARTING

Autotransformer starters provide reduced-voltage starting at the motor terminals through the use of a tapped, three-phase autotransformer. Upon initiation of the controller pilot device, a two- and a three-pole contactor close to connect the motor to the preselected autotransformer taps. A timing relay causes the transfer of the motor from the reduced-voltage start to line-voltage operation without disconnecting the motor from the power source. This is known as *closed transition starting.*

Taps on the autotransformer provide selection of 50%, 65%, or 80% of line voltage as a starting voltage. Starting torque will be 25%, 42%, or 64%, respectively, of line-voltage values. However, because of transformer action, the controller line current will be less than motor current, being 25%, 42%, or 64% of full-voltage values. This autotransformer starting may be used to provide maximum torque available with minimum line current, together with taps to permit both of these factors to be varied. Figure 12–19 shows torque and voltage tap points.

Manual autotransformer starters are used to start squirrel-cage polyphase motors when the characteristics of the driven load or power company limita-

FIGURE 12–19 *Autotransformer starting—speed versus torque. (The Lincoln Electric Co.)*



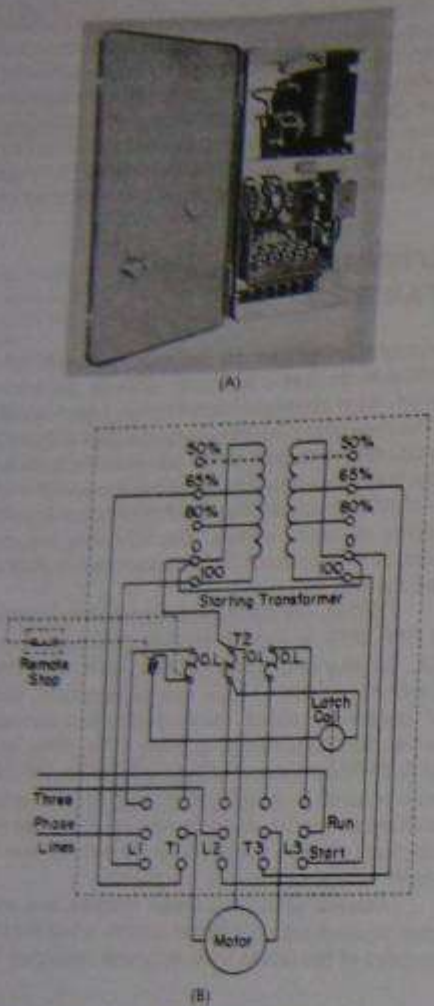


FIGURE 12-20 (A) Autotransformer type of magnetic starter, (B) corresponding wiring diagram. (Allen-Bradley)

tions require starting at reduced voltage (Fig. 12-20). NEMA (National Electrical Manufacturers Association) permits one start every 4 minutes, for a total of four starts followed by a rest period (2 hours). Each starting period is not to exceed 15 seconds. Figure 12-21 shows a autotransformer type of starter. Note the location of the taps on the starting transformer.

The autotransformer provides the highest starting torque per ampere of line current. Thus it is an effective means of motor starting for applications where the inrush current must be reduced with a minimum sacrifice of starting torque. This type of starter arrangement features closed-circuit transition, an arrangement that maintains a continuous power connection to the motor during the transition from reduced to full voltage. This avoids the high transient switching currents characteristic of starters using open-circuit transition. It provides smoother acceleration as well.

Operation

Operating an external START button or pilot device closes the neutral and start contactors, applying reduced voltage to the motor through the autotransformer. After a preset interval, the timer contacts drop out the neutral contactor, breaking the auto-transformer connection but leaving part of the windings connected to the motor as a series reactor. The RUN contactor then closes to short out this reactance and apply full voltage to the motor. Transition from reduced to full voltage is accomplished without opening the motor circuit.

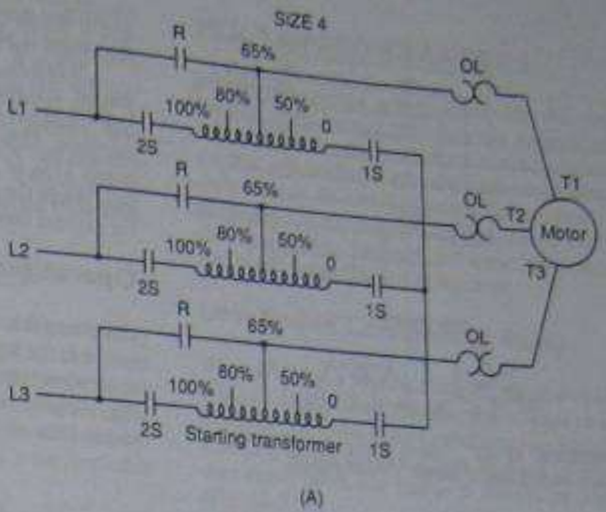
For starters rated up to 200 hp you should allow a 15-second operation out of every 4 minutes for 1 hour followed by a rest period of 2 hours. For starters rated above 200 hp, you should allow three 30-second operations separated by 30-second intervals followed by a rest period of 1 hour. The major disadvantages of this type of starter are its expense for lower horse-power ratings and its low power factor.

PART-WINDING STARTING

Part-winding motors have two sets of identical windings—intended to be operated in parallel—which can be energized in sequence to provide reduced starting current and reduced starting torque. Most (but not all) dual-voltage 230/460-V motors are suitable for part-winding starting at 230 V.

When one winding of a part-winding motor is energized, the torque produced is about 50% of "both winding" torque, and line current is 60 to 70% (depending on motor design) of comparable line voltage values. Thus, although part-winding starting is not truly a reduced voltage means, it is usually also classified as such because of its reduced current and torque.

When a dual-voltage delta-connected motor is operated at 230 V from a part-winding starter having a three-pole start and a three-pole run contactor, an unequal current division occurs during normal operation resulting in overloading of the starting contactor. To overcome this defect, some part-winding starters use a four-pole starting contactor and a two-pole run contactor. This arrangement eliminates the unequal current division obtained with a delta-wound motor, and it enables wye-connected part-winding motors to be given either a one-half or two-thirds part-winding start.



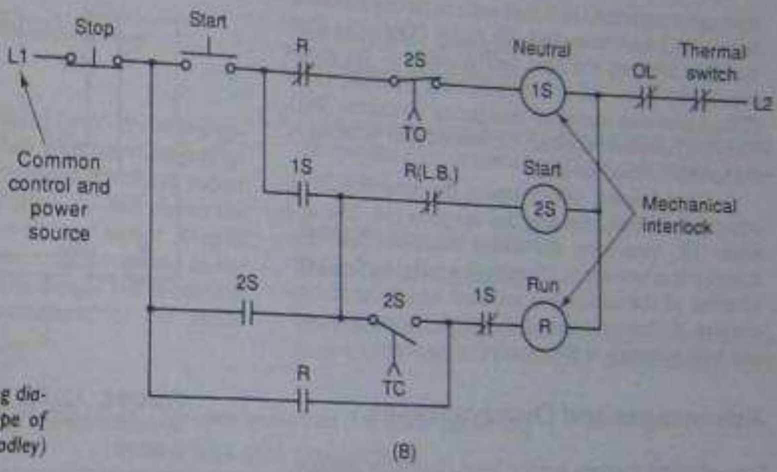


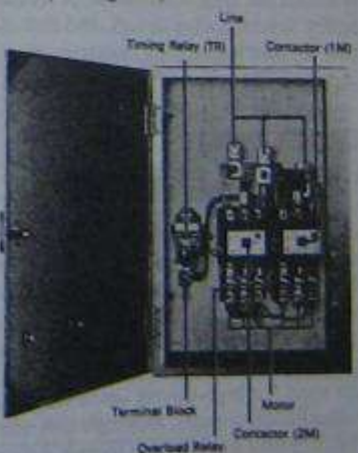
FIGURE 12-21 Typical wiring diagram for an autotransformer type of reduced-voltage starter. (Allen-Bradley)

The class 8640 starters have a start contactor, a timing relay, a run contactor, and necessary overload relays. Closing the pilot device contact causes the start contactor to close to connect the start winding and to initiate the time cycle. After expiration of the preset timing, the run contactor closes to connect the balance of the motor windings. A time setting of 1 second is recommended. Most motor manufacturers do not permit energization of the start winding alone for longer than 3 seconds. Part-winding starters provide closed transition starting.

Operation

The part-winding type of starter is shown in Fig. 12-22. The parts are located for ease in understanding the operation. By taking a look at the schematic in Fig. 12-23 you can see how the starter operates. Closing the START button or other pilot device energizes

FIGURE 12-22 Part-winding type of magnetic starter. (Westinghouse)



draw about 33% of its normal locked-rotor current. After an adjustable time interval, the motor is automatically connected in delta, applying full line voltage to the windings. In starters with open-circuit transition the motor is momentarily disconnected from the line during the transition from the wye to delta. With closed transition (Fig. 12-25) the motor remains connected to the line through the resistors. This avoids the current surges associated with open-circuit transition.

Advantages and Disadvantages

The advantages are moderate cost and its suitability for high-inertial, long-acceleration loads. It does have torque efficiency. However, the disadvantages are that it requires special motor design, starting torque is low, and it is inherently open transition—closed transition is available at added cost. There is no flexibility in selecting starting characteristics.

Star–Delta (Wye–Delta) Connections

There is the 12-lead motor wound for Y–Δ starting operation on either low voltage or a higher voltage (Fig. 12-26). There is also a six-lead single-voltage motor suitable for Y–Δ starting. Figure 12-26B shows the connection to the lines for the six-lead motor. Keep in mind that overload relay protection is required by the *National Electrical Code*®. The size of the protection is determined by the manufacturer of the motor (Table 12-4).

MULTISPEED STARTERS

Multispeed starters are designed for the automatic control of two-speed squirrel-cage motors of either the consequent pole or separate winding types. These starters are available for constant-horsepower, constant-torque, or variable-torque three-phase motors. Multispeed motor starters are commonly used on machine tools, fans, blowers, refrigeration compressors, and many other types of equipment.

Low-Speed Compelling Relay

When added to a standard starter, the low-speed compelling relay compels the operator always to start the motor in low speed before switching to a higher speed. This is a safety feature where damage to equipment may result when the motor is started at high speed (Fig. 12-27).

Automatic Sequence Accelerating Relay

The automatic sequence accelerating relay will control the sequence of acceleration from low speed up to high speed.

Automatic Sequence Decelerating Relay

The automatic sequence decelerating relay is used with large-inertia loads. The braking effect caused by a

FIGURE 12-26 *Star–delta connections. (The Lincoln Electric Co.)*

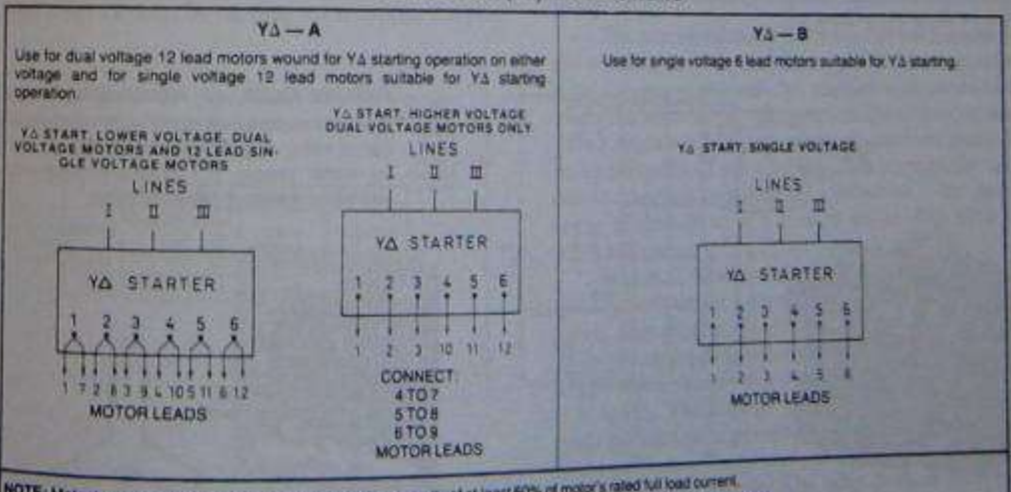
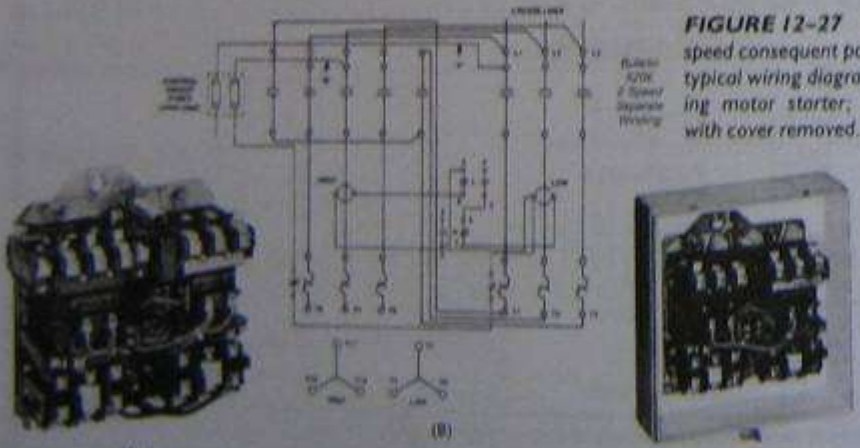


TABLE 12-4 SELECTION OF A CONTROLLER BEST SUITED FOR A PARTICULAR CHARACTERISTIC

Characteristic Wanted	*Type of Starter to Use (Listed in Order of Desirability)*	*Comments*
Smooth acceleration	1. Solid state (class 8660) 2. Primary resistor (class 8647) 3. Wye-delta (class 8630) 4. Autotransformer (class 8606) 5. Part-winding (class 8640)	Little choice between 3 and 4.
Minimum line current	1. Autotransformer (class 8606) 2. Solid state (class 8660) 3. Wye-delta (class 8630) 4. Part winding (class 8640) 5. Primary resistor (class 8647)	
High starting torque	1. Autotransformer (class 8606) 2. Solid state (class 8660) 3. Primary resistor (class 8647) 4. Part winding (class 8640) 5. Wye-delta (class 8630)	Little choice between 3, 4, and 5.
High torque efficiency (torque vs. line current)	1. Autotransformer (class 8606) 2. Wye-delta (class 8630) 3. Part winding (class 8640) 4. Solid state (class 8660) 5. Primary resistor (class 8647)	
Suitability for long acceleration	1. Wye-delta (class 8630) 2. Autotransformer (class 8606) 3. Solid state (class 8660) 4. Primary resistor (class 8647)	For acceleration time greater than 5 seconds, primary resistor requires non-standard resistors. Part-winding controllers are unsuitable for acceleration time greater than 3 seconds.
Suitability for frequent starting	1. Wye-delta (class 8630) 2. Solid state (class 8660) 3. Primary resistor (class 8647) 4. Autotransformer (class 8606)	Part-winding is unsuitable for frequent starts.
Flexibility in selecting starting characteristics	1. Solid state (class 8660) 2. Autotransformer (class 8606) 3. Primary resistor (class 8647)	For primary resistor, resistor change required to change starting characteristics. Starting characteristics cannot be changed for wye-delta or part-winding controllers.

Source: Courtesy of Square D.



(A) (B) (C)

FIGURE 12-27 *(A) Multispeed starter and two-speed consequent pole starter without enclosure; (B) typical wiring diagram for two-speed separate winding motor starter; (C) general-purpose enclosure with cover removed. (Allen-Bradley)* sudden change from high to low speed may cause damage to the motor or to the driven machine. To avoid this danger, the operation should give the motor sufficient time to slow down by pushing the STOP button and then waiting a short interval before pushing the button for a lower speed.

To help provide correct operation, multispeed starters can be equipped with an automatic sequence decelerating relay for each lower-speed step. This relay automatically interposes a time delay between the speed steps and makes it unnecessary to press the STOP button when switching to a lower speed.

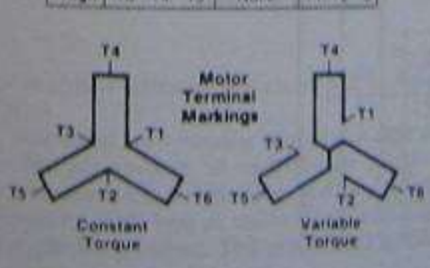
CONSEQUENT-POLE MOTOR CONTROLLER

By increasing the number of poles a motor has it is possible to change its speed. By increasing the number of poles, the speed of the motor is decreased. Inasmuch as a motor is wound and mounted rather permanently on a frame, it is not easily possible to take out or put in poles or the associated windings. Therefore, an electrical means must be found if the speed of the motor is to be changed by using the number of poles method to do so. One method of doing this is the consequent-pole arrangement. This method can be used for two-speed, one-winding motors or four-speed, two-winding motors.

The reversal of some of the currents in the windings has the same effect as physically increasing or decreasing the number of poles. Three-phase motors are wound, in some cases, with six leads brought out for connection purposes. It is possible to connect the windings, using combinations of the terminals for connection purposes, either in series delta or in parallel wye (Fig. 12–28). By tapping the windings it is possible to send current in two different directions, effectively creating more poles and decreasing the speed of the motor. The number of poles is doubled by reversing through half a phase. Two speeds are obtained by producing twice as many consequent poles for low-speed operation as for high speed.

FIGURE 12–28 Connections made by the consequent-pole starter for constant torque or variable torque. (Allen-Bradley)

CONNECTIONS MADE BY STARTER			
Speed	Supply Lines L1 L2 L3	Open	Together
Low	T1 T2 T3	T4 5 6	None
High	T6 T4 T5	None	T1 2 3



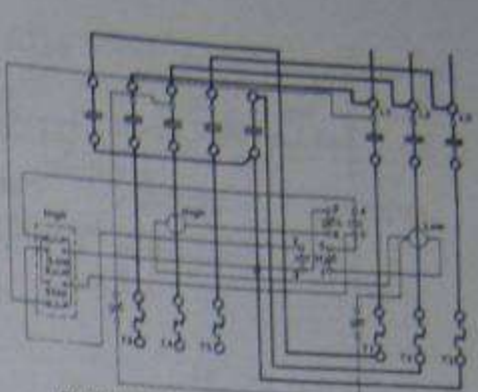


FIGURE 12–29 Wiring diagram for a two-speed, consequent-pole, constant-horsepower motor, NEMA size 0–4. (Square D)

Figure 12–29 shows how the controller is wired to produce consequent poles for constant torque or variable torque. The wiring diagram and the line drawing (Fig. 12–30) illustrate connections for the following method of operation: The motor can be started in either HIGH or LOW speed. The change from LOW to HIGH or from HIGH to LOW can be made without first pressing the STOP button. Figure 12–31 shows pilot devices with connections that can be made to obtain different sequences and methods of operation. The series delta arrangement produces high speed. It also produces the same horsepower rating at high and low speeds.

The torque rating is the same for both speeds if the winding is such that the series delta connection gives the low speed and the parallel wye connection gives the high speed. Consequent-pole motors that have a single winding for two speeds have the extra tap at the midpoint of the winding. This permits the various connection possibilities. However, the speed range is limited to a 1:2 ratio of or 600/1200 or 900/1800 rpm.

Figure 12–32 shows the motor terminal markings and connections for a constant-horsepower delta. The wiring diagram (Fig. 12–33) and the line drawing (Fig. 12–34) illustrate connections for the following method of operation: Motor can be started in either HIGH or LOW speed. The change from LOW to HIGH can be made without first pressing the STOP button. When changing from HIGH to LOW, the STOP button must be pressed between speeds. The pilot devices shown in Figure 12–35 show the other connections that can be

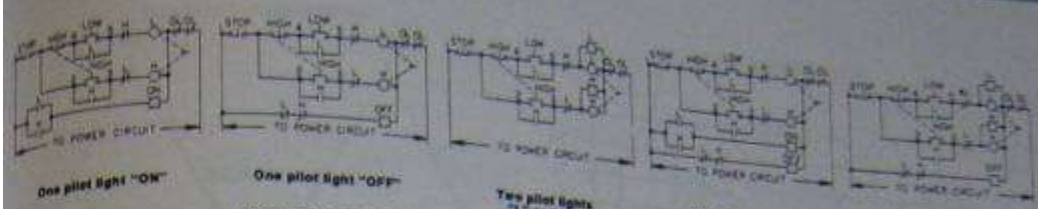


FIGURE 12-34 Elementary drawing of the control circuitry for a consequent-pole starter. (Allen-Bradley)

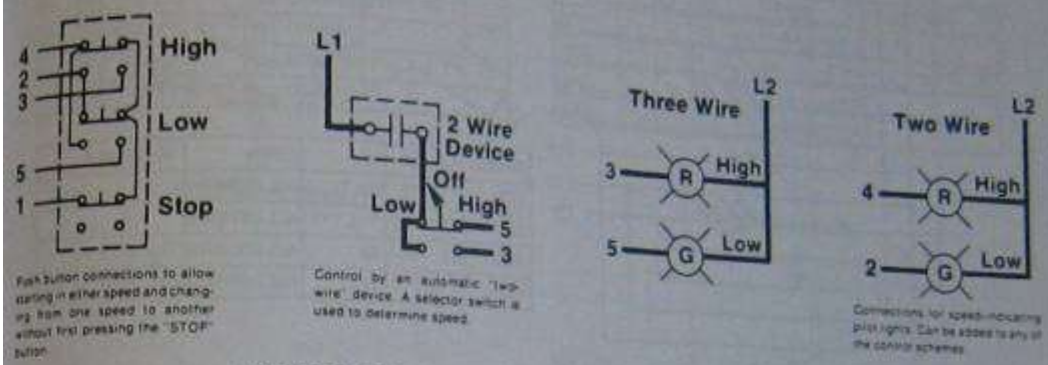


FIGURE 12-35 Connections for different sequences and methods of operation. (Allen-Bradley)

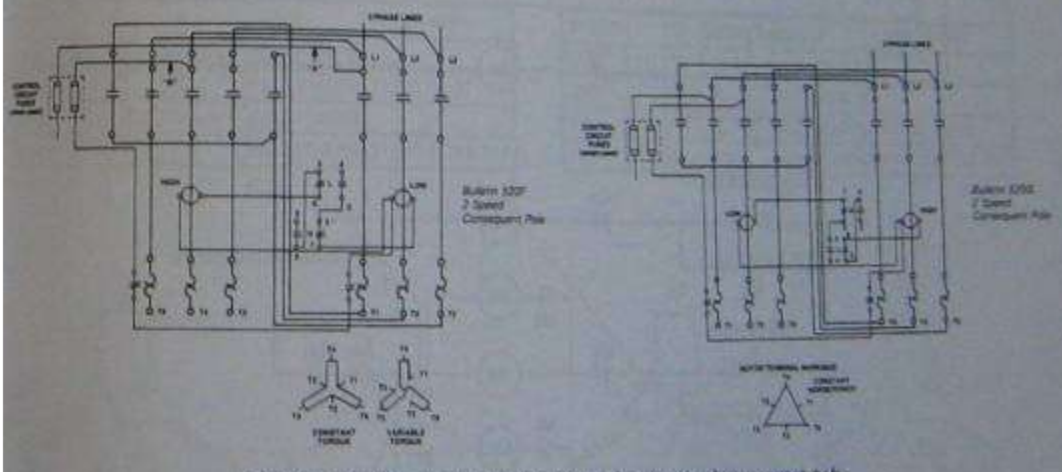
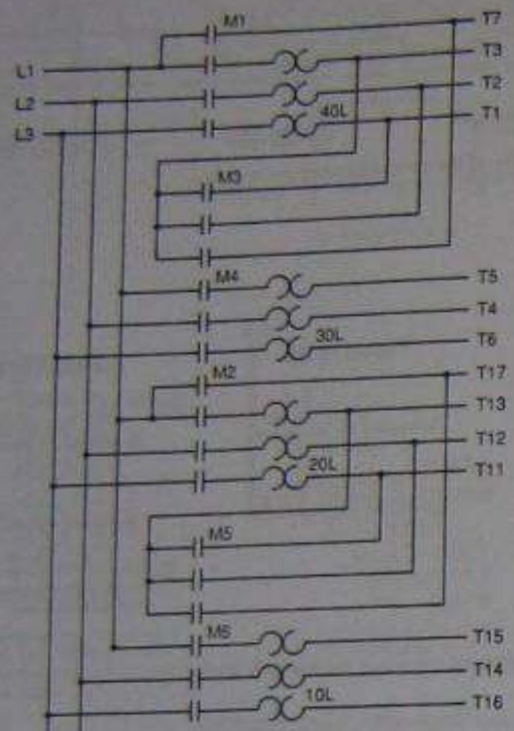


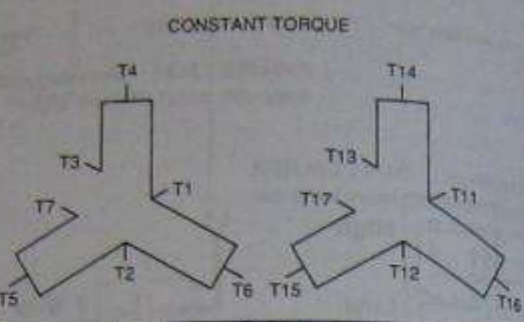
FIGURE 12-36 Typical wiring diagrams for two-speed consequent-pole starter. (Allen-Bradley)

made to obtain different sequences and methods of operation.

Four-speed, two-winding consequent-pole motor controllers can be used on squirrel-cage motors that have two reconnectable windings and two speeds for each winding. This type of motor does need a special type of starting sequence. This means that it must use the properties of the compelling relay, accelerating relay, and decelerating relay to operate correctly.

Figure 12-36 shows the two-speed consequent-pole starter with variable-torque and constant-torque connections. Figure 12-37 shows how the four-speed, two-winding controller is connected for the possible arrangements using this type of motor.





SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3, T7	ALL OTHERS	
2ND	T11	T12	T13, T17	ALL OTHERS	
3RD	T6	T4	T5	ALL OTHERS	T1, T2, T3, T7
HIGH	T16	T14	T15	ALL OTHERS	T11, T12, T13, T17

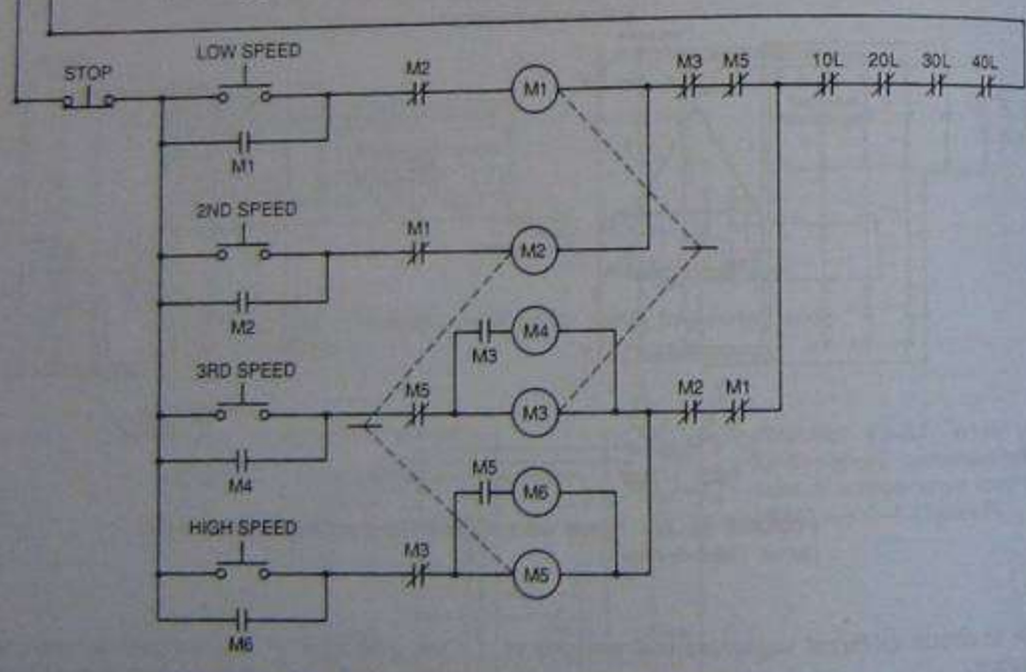
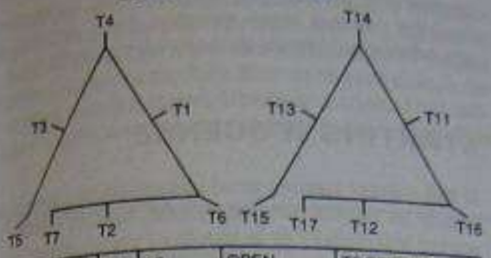


FIGURE 12-37 *Elementary diagram of a four-speed, two-winding controller and the possible arrangements for motor connections. (Allen-Bradley)*

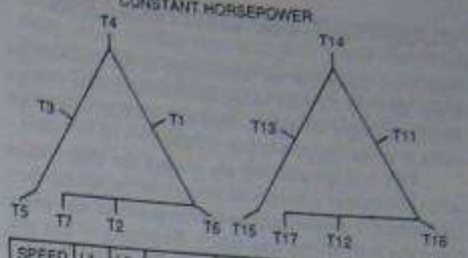
CONSTANT HORSEPOWER



SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3	ALL OTHERS	T4, T5, T6, T7
2ND	T6	T4	T5, T7	ALL OTHERS	
3RD	T11	T12	T13	ALL OTHERS	T14, T15, T16, T17
HIGH	T16	T14	T15, T17	ALL OTHERS	

(A)

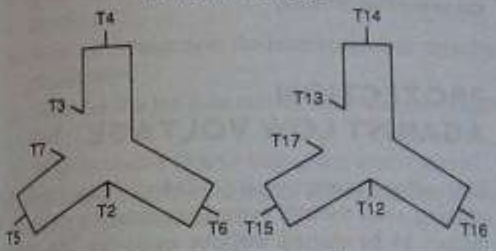
CONSTANT HORSEPOWER



SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3	ALL OTHERS	T4, T5, T6, T7
2ND	T11	T12	T13	ALL OTHERS	T14, T15, T16, T17
3RD	T6	T4	T5, T7	ALL OTHERS	
HIGH	T16	T14	T15, T17	ALL OTHERS	

(B)

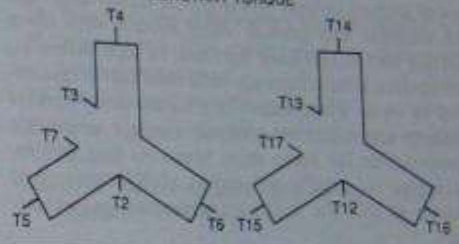
CONSTANT TORQUE



SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3, T7	ALL OTHERS	
2ND	T6	T4	T5	ALL OTHERS	T1, T2, T3, T7
3RD	T11	T12	T13, T17	ALL OTHERS	
HIGH	T16	T14	T15	ALL OTHERS	T11, T12, T13, T17

(C)

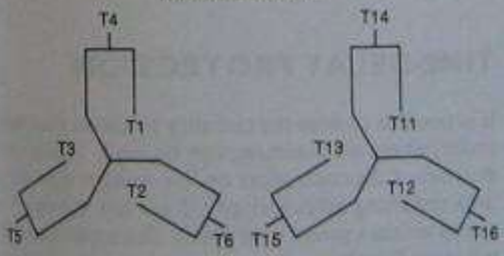
CONSTANT TORQUE



SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3, T7	ALL OTHERS	
2ND	T11	T12	T13, T17	ALL OTHERS	
3RD	T6	T4	T5	ALL OTHERS	T1, T2, T3, T7
HIGH	T16	T14	T15	ALL OTHERS	T11, T12, T13, T17

(D)

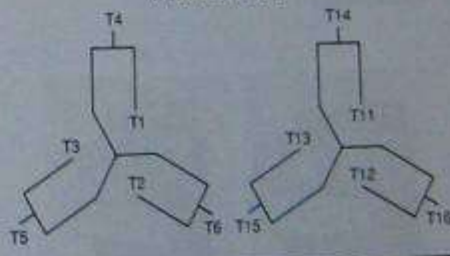
VARIABLE TORQUE



SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3	ALL OTHERS	
2ND	T6	T4	T5	ALL OTHERS	T1, T2, T3
3RD	T11	T12	T13	ALL OTHERS	
HIGH	T16	T14	T15	ALL OTHERS	T11, T12, T13

(E)

VARIABLE TORQUE



SPEED	L1	L2	L3	OPEN	TOGETHER
LOW	T1	T2	T3	ALL OTHERS	
2ND	T11	T12	T13	ALL OTHERS	
3RD	T6	T4	T5	ALL OTHERS	T1, T2, T3
HIGH	T16	T14	T15	ALL OTHERS	T11, T12, T13

(F)

FIGURE 12-37 *continued*

FULL-VOLTAGE CONTROLLERS

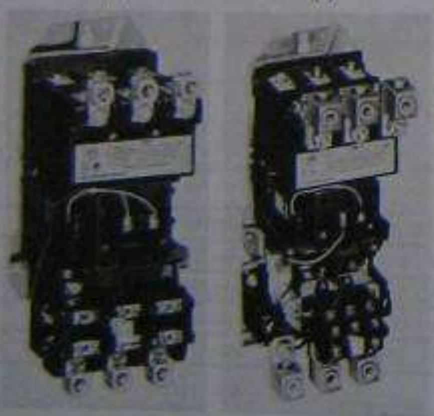
The least expensive of the starters is the full-voltage type. There is no limit to the horsepower, size, voltage rating, or type of motor that can be started on full voltage when the power is available.

Full-voltage starters are always the first choice *when the power system can supply initial inrush current*, and the motor and the driven machine can withstand the sudden starting shock. Examples of this are machines that start unloaded, as well as those that require little torque; or machines may be equipped with some form of unloading device to reduce starting torque, as in the use of an unloader valve in a compressor. A clutch may be inserted between a machine and motor so that the motor may be started unloaded. When the motor is up to speed the clutch is engaged. Clutches are sometimes used on large machines so that maximum horsepower can be exerted during breakaway without serious power system disturbance. Use of clutches also permits using motors with lower torque and locked-rotor currents. In most instances, up-to-date installations use *solid-state motor controllers* to better advantage. Many of the older types of starters are still in use and will continue to provide good service for many more years. As they deteriorate, they are usually replaced by a solid-state type of starter so that the clutch arrangements are unnecessary.

Figure 12-38 shows the general-purpose enclosure for a full-voltage starter. This type of starter is designed for full-voltage starting of polyphase squirrel-cage motors and primary control of slip-ring motors. This type of starter may be operated by remote control with pushbuttons, float switches, thermostats, pressure switches, snap switches, limit switches, or any other suitable two- or three-wire pilot device.

FIGURE 12-38 *Full-voltage starters (NEMA), open type, without enclosure: (A) size 3; (B) size 5. (Allen-Bradley)*

(A) (B)



STARTING SEQUENCE

If full-voltage starting produces excessive current demands on the distribution system, motors should be started individually or in blocks of permissible size by using some method of time delay, such as motor driver, pneumatic, or mercury plunger timing relays. When large and small motors are to be started on a common power system, best results are obtained by starting the largest sizes first. This gives larger motors the advantage of full-line capacity. If synchronous motors are on the system with other types of ac motors, the synchronous units should always be started first since they provide voltage stability for starting the induction motors.

PROTECTION AGAINST LOW VOLTAGE

Low-voltage protection is needed while the motors are running even though systematic starting permits all motors to be started without excessive line voltage drop. When three-wire control circuits are used, a severe dip in line voltage or a momentary complete outage breaks the control-sealing circuits, and the controller drops out and stops the motor. This provides low-voltage protection and prevents simultaneous acceleration of all motors to full speed after being slowed down by a voltage dip. However, all motors are disconnected from the line during the voltage dip, and each must be restarted.

TIME-DELAY PROTECTION

It is possible to wire the circuitry so that a time-delay undervoltage arrangement can be used. This permits dropout of the controllers on low-voltage dips but allows restarting automatically if normal voltage is restored within a preset time delay. The usual time delay is 2 seconds or less.

Time-delay undervoltage protection on controllers will prevent some complete shutdowns but should be applied with caution. If used on all motor controllers, restoration of voltage within the time-delay setting after a voltage dip causes each motor to attempt to accelerate simultaneously, thus producing excessive currents that may operate backup protection and starter overload devices and disconnect the motors.

Pilot devices such as pressure, float, or temperature switches automatically start and stop motors as the demand arises. On severe voltage dips or voltage failure, motor controllers drop open even though the demand switch is closed. Upon restoration of full voltage all units attempt to restart at the same time. This operating hazard can be overcome by adding a time delay in the starting circuit of each motor and timing the demand for starting at slightly different intervals. Time delays of various units can then be staggered so that at the restoration of voltage only one unit at a time will be started.

QUESTIONS

1. What is voltage spread?
2. What is the purpose of a centrifugal switch on a single-phase motor?
3. How can direction of rotation be reversed on a split-phase motor?
4. What type of motor uses pushrods and a wound armature?
5. Where are capacitor-start motors used?
6. How are capacitor-start motors reversed when standing still?
7. What advantage does the permanent-split capacitor motor have?
8. What are shaded-pole motors most likely to be used for?
9. What is needed to get a split-phase motor to run?
10. How much current does an across-the-line motor draw when it starts?
11. What is the advantage of reduced-voltage motor starting?
12. What is another name for primary resistor starters?
13. What is the major disadvantage of the autotransformer starter?
14. What type of starting does part-winding starters provide?
15. What is the least expensive method of motor starting?
16. Where are wye–delta starters typically used?
17. Why are wye–delta starters used with delta-wound squirrel-cage motors?
18. Why are compelling relays needed?
19. What happens to motor speed when more poles are added?
20. How do consequence pole motors obtain two speeds?

Solid-State Reduced-Voltage Starters

Objectives

After studying this chapter, you will be able to:

1. Define thyristor operation.
2. Explain what gating does.
3. Explain solid-state stepless acceleration.
4. Describe the operation of a diac in a control circuit.
5. Describe the operation of a triac in a control circuit.
6. Explain how surge supressors are installed on magnetic devices.
7. Describe lightning surge protection.

The electromechanical devices used for years are still reliable and working in many installations. They are used to provide sequencing and interlocking tasks. They are simple in construction, flexible in use, and have many contact combinations. They can also handle large currents and break the circuit as required.

Solid-state devices have no moving parts and no contacts to clean, replace, or adjust. They use transistors, triacs, diacs, and SCRs to do the switching. These logic elements can perform the same functions in a solid-state system as relays do in the electromechanical systems (Fig. 13-1).

The solid-state control device has many advantages that make it desirable for the various environments in which it has to operate. It has no contacts to become dirty or malfunction when needed to control a critical sequence of operations. The solid-state control devices are more reliable than electromechanical devices. They come in sealed-in modules that can be plugged into a rack and replaced as a unit if anything goes wrong with the circuitry.

FIGURE 13-1 *Solid-state reduced-voltage controller. (Square D)*

REDUCED-VOLTAGE STARTING

Reduced-voltage starting can be accomplished in a number of ways. However, in solid-state circuitry it is somewhat simpler than described previously. The exact details of the circuit functions are somewhat more complex than those of the electromechanical system; however, a complete understanding of solid-state physics and/or electronics is not necessary in order to grasp the workings of the simple devices utilized to perform the operations of solid-state switching and control.

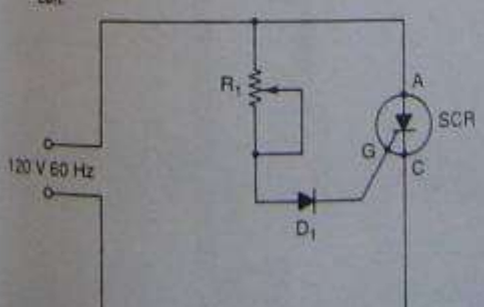
SILICON-CONTROLLED RECTIFIERS

The silicon-controlled rectifier (SCR) is the device used most often to control electric motors. The proper name for an SCR is *thyristor*. However, popular use of the term SCR has made it part of the literature and accepted by everyone working in the field. It is a specialized type of semiconductor used for control of electrical circuits.

An SCR conducts current in a forward direction only. The symbol for an SCR is shown in Fig. 13-2. Current flows through an SCR from the cathode (C) to the anode (A). The illustration indicates the SCR also has a gate (G).

FIGURE 13-2 *Symbol for SCR.*

FIGURE 13-3 *Schematic of SCR-controlled circuit*



The function of the SCR is shown in the circuit diagram in Fig. 13-3. The most typical use of an SCR is for a controlled circuit. Examples include a light dimmer or a speed control for a motor. This type of circuit is illustrated in Fig. 13-3. The resistor in this circuit, R_1, is a rheostat, or adjustable resistor. This is used to control the amount of voltage delivered to the gate of the SCR. The more voltage delivered, the greater the flow. Thus, adjusting the rheostat can serve to control the circuit. If the circuit illuminates a lamp, lowering the voltage to the rheostat dims the bulb. If the load is a motor, its speed is slowed. Figures 13-4, 13-5, and 13-6 show what typical SCRs look like with their leads identified according to cathode, gate, and anode connections.

One of the main reasons for using semiconductor devices for motor control is the device's ability to start a motor under reduced-voltage conditions and thus allow the motor to accelerate to full speed at a lower torque level. By reducing the high current inrush the mechanical shock to the driven equipment is reduced.

A reduced-voltage solid-state motor starter uses SCRs for power control. Inasmuch as an SCR conducts in the direction of the arrow in the symbol, it

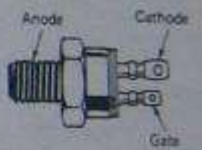


FIGURE 13-4 *Drawing of a typical SCR.*

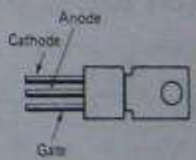
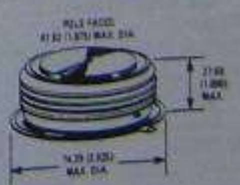


FIGURE 13-5 *Drawing of typical SCR.*

FIGURE 13-6 *Larger currents require larger SCRs.*



means that current flows only one way in an SCR. To use an SCR to its advantage on ac it is necessary to use two of them in reverse parallel (Fig. 13-7). SCRs have to be turned on in order to conduct current through them; that is, they need a gate pulse to turn them on. Once an SCR is turned on or gated, it does not stop forward current flow. Full wave control uses two

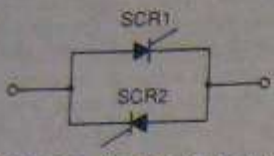


FIGURE 13-7 *Parallel SCRs for one phase.*

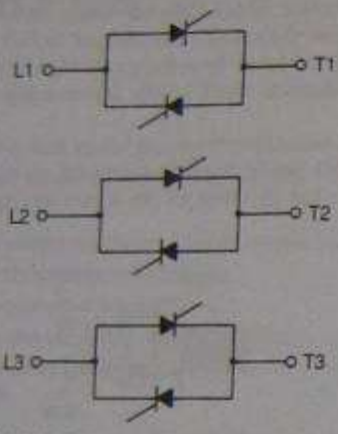
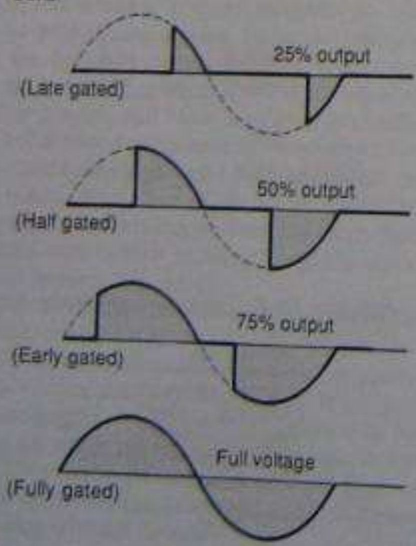


FIGURE 13-8 *Three-phase SCR arrangement.*

FIGURE 13-9 *Outputs from differently gated SCRs.*



SCRs in *each phase*. Three-phase operation mu lize six diodes, connected as shown in Fig. 13-8.

The current through an SCR can be contr by *gating* the SCR at different times within the cycle. This also controls the acceleration time of motor. If the gate pulse is applied early in the cycle, the output is high. If the gate pulse is app late in the half-cycle, only a small part of the w form is passed through and the output is low. S controlling the SCR's output voltage the motor ac eration characteristics can be controlled (Fig. 13-

SOLID-STATE STEPLESS ACCELERATION

The class 8660 solid-state reduced-voltage contro provides smooth, stepless acceleration of a thr phase induction motor. The controller offers seve standard and option features to control, monitor, protect the motor during the start and run modes operation. Modular construction of the control adds flexibility and ease of maintenance (Fig. 13-1 Soft start is accomplished by gradually turning on silicon-controlled rectifiers. Two SCRs are connec in a back-to-back or reverse-parallel arrangement a mounted on a heat sink to make up a power pole. T power pole also contains a printed circuit board and thermal sensor.

Firing of the SCRs is controlled by the modul on the logic rack. These modules also check for co rect startup and running conditions and provide a vi ual indication of controller status through the use light-emitting diodes (LEDs). Each module has a sp

FIGURE 13-10 *Power pole. (Square D)*

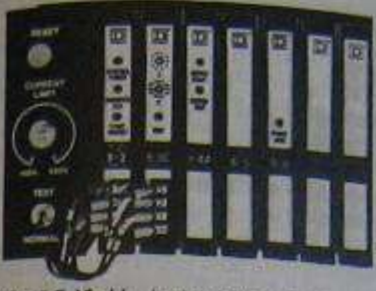


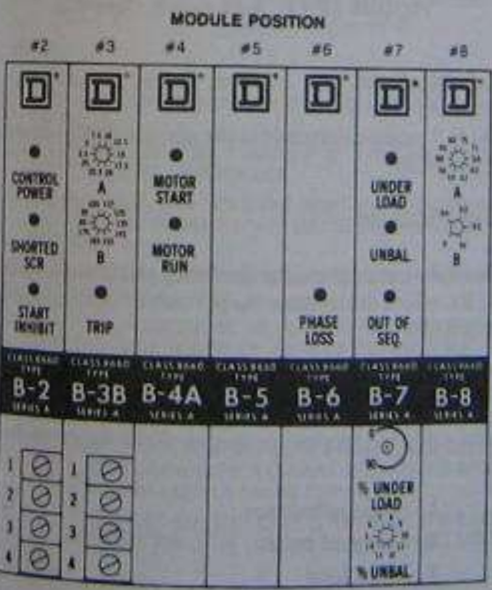
FIGURE 13–11 Logic module rack. (Square D)

cific location and function. Figure 13–11 shows a logic module rack.

LOGIC RACK

The logic rack is located on the lower part of the controller and has sockets for eight plug-in modules (Fig. 13-12). Each module has a specific location and performs a specific function in the operation of the controller. The module in the first position is internal to the controller and provides wiring connections between the power pole and the logic modules. The modules in positions 2 through 8 control the firing of the SCRs, check for correct startup and running conditions, and provide a visual indication of the controller status through the use of LEDs. The B-2 module goes in the second position, one of the B-3 modules goes in

FIGURE 13–12 Module position in logic rack. (Square D)



the third position, and so on. The specific module functions are described below.

B-2 Module

This module provides logic voltages and checks for correct starting conditions. The control can be started if the control power LED is ON and the start inhibit LED is OFF.

B-3 Module

A three-phase, temperature-compensated, solid-state overload relay is supplied as an integral part of the controller. It provides class 10, inverse-time trip characteristics that protect against harmful motor overloads. There is a different B-3 module for each of the four controller current ratings of 200, 320, 500, and 720 A. Motor full-load current settings are adjustable by the use of potentiometers on the B-3 module. An overload condition will automatically deenergize the controller, close the alarm contact, and light the TRIP and START INHIBIT LEDs. An overload test feature on the logic rack assembly provides a check for operation of the solid-state overload circuitry. Overload trip time is a function of the current limit setting. The lower the current limit setting, the longer the trip time. Trip times for three current limit settings are shown in Table 13-1. Longer trip times for high-inertial loads can be provided on the other types of controllers. Form Z72 provides class 30 inverse-time trip characteristics by using a special B-3 module and power poles with higher current ratings. Trip times for class 30 overloads are shown in Table 13-1.

B-4 Module

The starting method that is used is determined by the B-4 module. Current limit starting is standard. The current limit setting is adjustable by the use of a potentiometer on the logic rack assembly. Optional starting methods are available. A description of each of the starting methods follows.

Current Limit (B-4A Module). The current-limit feature will limit the motor current to a preset level at all times during start and run conditions. Current limit

TABLE 13–1 OVERLOAD TRIP TIMES

Current Limit (% of MFLC)	Trip Time (seconds)	
	Standard Class 10	Form Z72 Class 30
150	90	250
300	30	90
425	5	40

is adjustable between 150 and 425% of motor full-load current by way of a potentiometer located on the logic rack. If a shorting contactor is used, this feature will be present only in the start condition (Fig. 13–13).

Linear Timed Acceleration (B-4B) Tachometer Feedback. This option allows the motor speed to be increased linearly with the time until the motor reaches full speed (Fig. 13–14). Start time is adjustable from 3 to 30 seconds and does not fluctuate with motor loading. This method gives the smoothest acceleration but requires a tachometer input. Motor current is limited to the current-limit setting.

Voltage Ramp (B-4C Module). This option allows the applied motor voltage to increase linearly from 0 to 100% over an adjustable period of 3 to 30 seconds. The motor current is limited to the current-limit setting. This method provides acceleration that is approximately linear from zero to full speed but does not require a tachometer. The actual acceleration time depends on the motor and load (Fig. 13–15).

Current Ramp (B-4D Module). This option supplies a breakaway current to the motor at start and then linearly ramps the current up to the current-limit setting. Breakaway current is adjustable from 0 to 150% of motor full-load current. Ramp time is adjustable from 0 to 7 seconds. This method provides the greatest control of starting current (Fig. 13–16).

Accel/Decel (B-4E Module). This option provides both a soft start and a soft stop. Starting characteristics are identical to the voltage ramp start (B-4C). This option also allows the applied motor voltage to decrease linearly from 50% to 0% over an adjustable period of 3 to 30 seconds to provide a soft stop. Provisions for an emergency stop are included (Fig. 13–17).

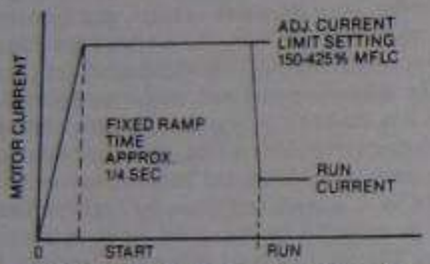


FIGURE 13–13 *B-4A module. (Square D)*

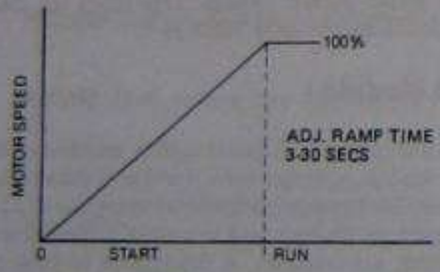


FIGURE 13–14 *B-4B module. (Square D)*

FIGURE 13–15 *B-4C module. (Square D)*

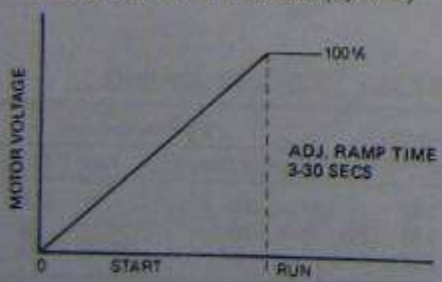


FIGURE 13–16 *B-4D module. (Square D)*

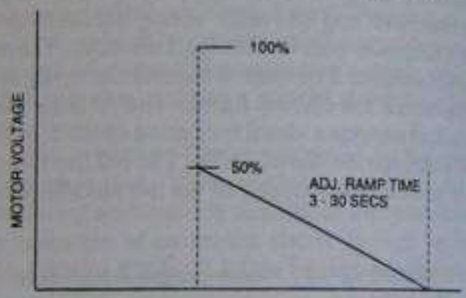


FIGURE 13–17 *B-4E module. (Square D)*

B-5 Module

The B-5 module determines the correct firing sequence of the SCRs.

B-6 Module

The B-6 module provides the firing phase angles of the SCRs, which determines the percent of conduction for each SCR.

B-7 Voltage Monitor Module

This optional module provides three separate functions:

1. Phase unbalance
2. Phase reversal
3. Underload

If any one of these occurs, the controller will shut off and the appropriate LEDs will be lighted.

The *phase unbalance function* is activated whenever three-phase power is present at the controller line terminals but is disabled during starting. A fault condition occurs when voltage unbalance is greater than the unbalance setting. The voltage unbalance setting is adjustable from 5% to 14% as defined by NEMA standards.

The *phase-reversal function* is activated whenever three-phase power is present at the controller line terminals. A fault condition occurs if the three phases are not in correct sequence. Without the B-7 module, the controller is phase insensitive and will operate with any phase sequence.

The *underload function* is activated after the motor is "up to speed." A fault condition occurs when the motor drops below the underload setting, which is adjustable from 0 to 90% of motor full-load current. This can be disabled by adjusting the setting to zero.

B-8 Energy-Saving Module

The energy-saving module will automatically adjust the voltage to the motor when load fluctuations occur. The motor will maintain full speed and required torque but draw less kVA when the load decreases. If the load increases, the module will respond by increasing the kVA so that the motor and load do not slow down in speed. This feature cannot be used on controllers with shorting contactors.

SHORTED SCR SWITCH

If an SCR shorts, the short is detected and the shorted SCR switch will flip to the YES position. This switch will also trip the shunt trip circuit breaker (if used) ahead of the controller. If there is an open circuit between the controller and the motor, the shorted SCR circuitry will trip. This can occur if there is an open disconnect switch between the controller and motor. Isolation contactors should be placed ahead of the controller. A motor load must be connected to the controller to prevent nuisance tripping of the shorted SCR circuitry.

ELEMENTARY WIRING DIAGRAMS FOR SOLID STATE

The solid-state reduced-voltage controller with an isolation contactor is shown in Fig. 13–18. Keep in mind that the M, SR2, OT, ALARM, SHORTED SCR, and UP-TO-SPEED relays are mounted on the controller and wired internally. Figure 13–19 shows the solid-state reduced-voltage controller with a shorting contactor and Fig. 13-20 shows the controller with a shorting contactor and an isolation contactor.

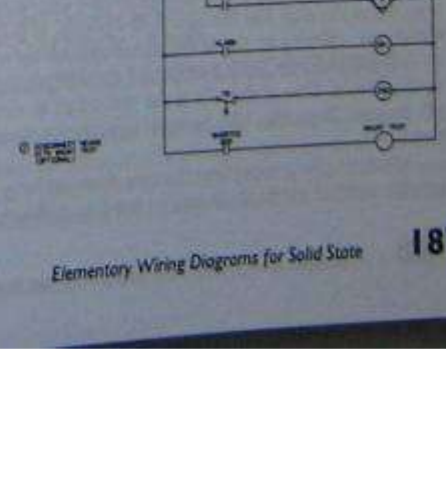
DIAC

The diac is basically a two-terminal device. It has a parallel-inverse combination of semiconductor layers.

FIGURE 13–18 Solid-state reduced-voltage controller with an isolation contactor. (Square D)

NOTES:

1. M, SR2, OT, ALARM, SHORTED SCR, AND UP-TO-SPEED RELAYS ARE MOUNTED ON THE CONTROLLER AND WIRED INTERNALLY.
2. M DENOTES THE COIL FUNCTION OF THE SOLID STATE REDUCED VOLTAGE CONTROLLER.
3. THE SR2 RELAY CONTROLS THE START AND STOP SEQUENCE, AND ALSO HAS CONTACTS THAT MAY BE USED AS ELECTRICAL INTERLOCKS.
4. OT IS AN OVER TEMPERATURE SWITCH THAT OPENS WHEN THAT CONDITION EXISTS.
5. OL IS THE OVERLOAD RELAY CONTACT. IT OPENS WHEN: AN OVERLOAD IS DETECTED; L1, L2 OR L3 VOLTAGE IS NOT PRESENT; OR THE 120V CONTROL VOLTAGE IS MISSING.
6. THE ALARM CONTACT CLOSES WHEN AN OVERLOAD IS DETECTED.
7. THE SHORTED SCR CONTACT CLOSES WHEN THAT CONDITION EXISTS. IT IS USED WITH A CIRCUIT BREAKER OR DISCONNECTING SWITCH WITH A SHUNT TRIP COIL.
8. THE UP-TO-SPEED CONTACT CLOSES WHEN THE SCR'S ARE IN FULL CONDUCTION. IT IS USED WITH A SHORTING CONTACTOR.



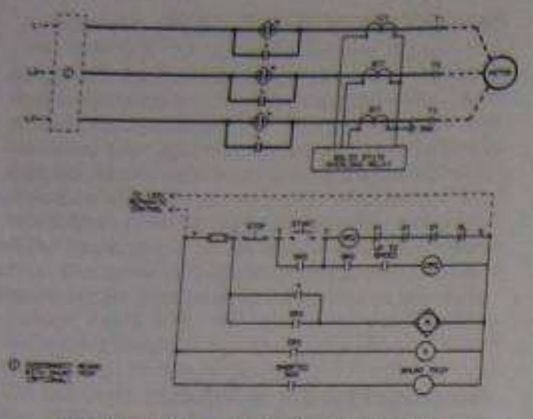


FIGURE 13-19 *Solid-state reduced-voltage controller with a shorting contactor. (Square D)*

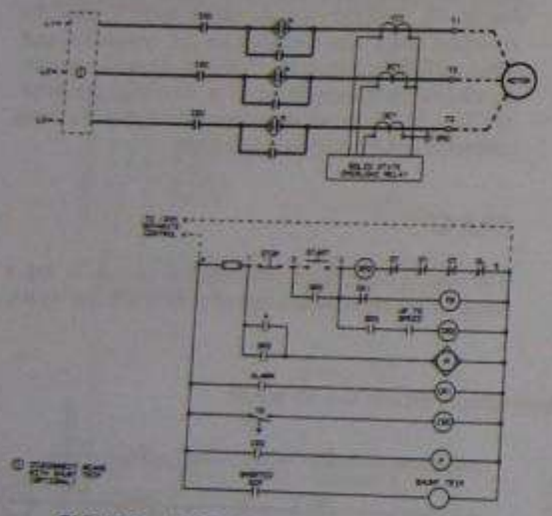


FIGURE 13-20 *Solid-state reduced-voltage controller with a shorting contactor and an isolation contactor. (Square D)*

This combination of layers permits the triggering of the device in either direction (Fig. 13-21). As you remember, the SCR allowed triggering in only one direction. Thus the diac has the ability to conduct in both directions when an ac signal voltage is applied across its terminals. There are a number of applications for such a device. One of them is in the control of ac electric motors. They may also be used in proximity detectors.

Note in the symbol that the diac does not have a gate or control element. It can be used as a bidirectional trigger diode (Fig. 13-23B). Current can flow either way when enough voltage is supplied for breakover. Typically, the firing potential is about 30 V in either direction. The diac is in its OFF state until the voltage across terminals T1 and T2 exceeds the breakover voltage. In power control circuits a diac can be used for more effective control of the *turn-on* point for the gate electrode of either a triac or an SCR.

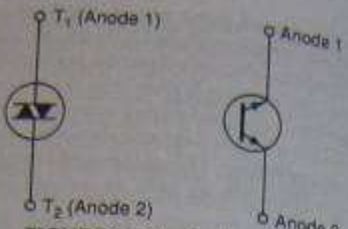


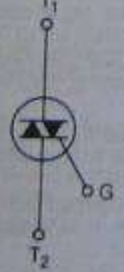
FIGURE 13-21 *Symbols for a diac.*

TRIAC

The triac is basically a diac with a gate terminal. The gate terminal controls the *turn-on* conditions of this bilateral device. The gate current can control the action of the device in either direction. This is similar to that of the SCR. However, the characteristics of the triac are somewhat different from those of the diac. Figure 13-22 shows the symbol and the location of the gate terminal.

By placing the triac in a circuit it is possible to indicate how it works (Fig. 13-23). In this arrangement the switch is used to select various conditions for the triac. The load can be either a light bulb or an ac motor. When the switch is in position 1 there is no gate connection. The triac does not conduct. The motor does not run. There is no trigger voltage applied to the gate. In position 2 a diode is placed in the circuit and with its polarity so arranged to allow a trigger voltage applied to the gate on the positive pulse of the ac applied to the circuit. The triac conducts, but only dur-

FIGURE 13-22 *Symbol for a triac.*



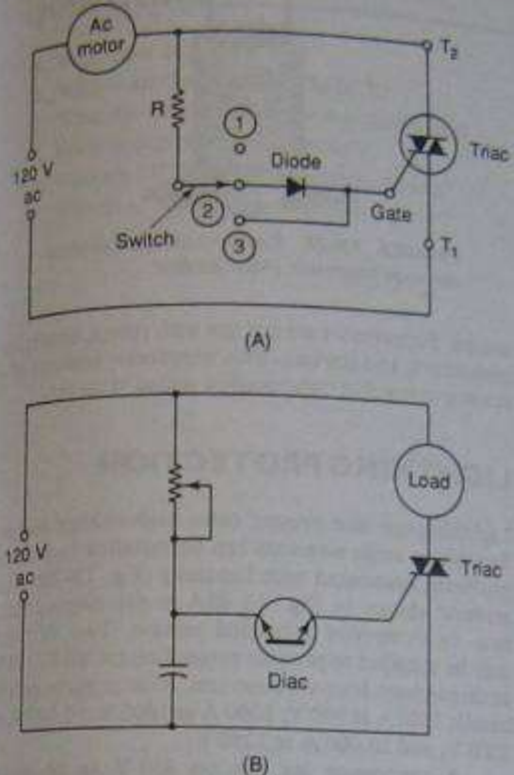


FIGURE 13-23 *(A) Triac demonstration circuit; (B) triac using a diac to trigger the gate.*

ing on one-half of the ac sine wave. This means that only about one-half of the normal current is applied to the motor. This is the same arrangement as with an SCR. An ac motor may have a problem with this type of pulsating dc voltage. When the switch is moved to position 3, the full ac sine-wave voltage is applied to the gate, with, of course, a reduction in value caused by the resistor R. Now that both halves of the ac sine wave are applied to the gate, the triac conducts full time and the full value of ac is applied to the ac motor. The motor then runs at full speed. R can be made a variable type and its value would then control the amount of ac current that passes through the triac and to the motor.

Another arrangement for the triac is shown in Fig. 13-23B. Here a diac is used to trigger the triac. The trigger voltage is controlled by the variable resistor. This allows for better regulation of the motor.

Triacs are packaged in the same types of cases as SCRs, so it is difficult or impossible to tell by a visual inspection which type is in the package. The numbers on the package indicate whether it is an SCR or a triac. There are triacs available today that can handle in excess of 10-kW loads.

LIGHT-EMITTING DIODES

Light-emitting diodes (LEDs) are used as indicator lights on the module panels for solid-state controllers. They are small, give off enough light for the purpose, and draw very little current. They are available in red, green, and amber (Fig. 13-24).

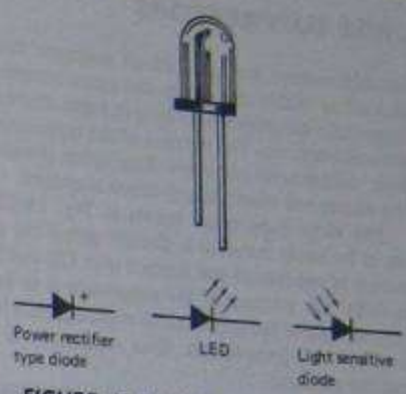


FIGURE 13-24 *Light-emitting diode (LED); symbol for LED.*

LEDs are made of gallium-arsenide junctions, a semiconductor material. Creation of electron-hole pairs is a reversible process. Energy is released when an electron recombines with a hole. In gallium-arsenide, an electron drops directly into a hole and a photon of energy is emitted. The gallium-arsenide junctions provide the best conditions for the generation of radiation in the visible range. Some are made for infrared radiation.

LEDs are used as indicator lamps. In most instances, they must be used in series with a resistor. They are also used as logic indicators for computer circuits. When reverse biased, the LED is nonconducting. This means that you have to have the proper polarity connections to the cathode and anode in order for it to glow. It is capable of conducting current when it is forward biased. It emits light when conducting with a forward bias current. An LED usually operates on 1 to 3 V. Excessive current will destroy an LED, and this calls for a series resistor in most circuits.

USING SOLID-STATE CONTROL AND ELECTROMAGNETIC DEVICES

When solid-state controls are utilized in circuits that have electromagnetic devices, there are problems with the "dirty" power source. The buildup and collapse of a magnetic field whenever a coil of wire or inductor is energized and deenergized produces spikes and other types of electrical *noise*. These spikes can cause problems with solid-state devices since they are susceptible to voltage surges and spikes that are commonplace with the energizing of relay coils and the turning on and off of electric motors.

SURGE SUPPRESSORS

Surge suppressors are installed on magnetic device coils, such as relays, contactors, and motor starters. A voltage-surge suppressor may have its leads connected to the coil terminals. The purpose of the suppressor is to limit voltage noise and overvoltage spikes produced by the starter coil when the coil circuit is opened.

The surge suppressor shown in Fig. 13-25 is made to be easily mountable directly across the coil terminals of contactors and starters with 120 and 240 V ac coils. The purpose of the suppressor is to limit voltage transients for applications requiring interface with solid-state components. One suppressor is required for each coil.

Figure 13-26 shows two types of surge suppressors used to reduce the high transient voltages generated when the coil circuit is opened. These suppressors are used with relay coils and other electromechanical devices.

Figure 13-27 is a surge suppressor used to protect solid-state devices against electrical transients that can result whenever electromechanical devices are operated. Suppressors are for use with relays, timers, contactors, and starters. This suppressor consists of a resistor–capacitor combination sealed in epoxy.

FIGURE 13-25 *Surge suppressor for mounting across coil terminals. (Allen-Bradley)*

FIGURE 13-26 *Surge suppressor: (A) for mounting under a relay; (B) for mounting on coil terminals. (Allen-Bradley)*

(A) (B)

FIGURE 13-27 *Resistor–capacitor combination surge suppressor. (Allen-Bradley)*

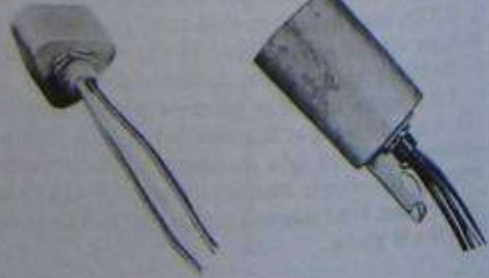
LIGHTNING PROTECTION

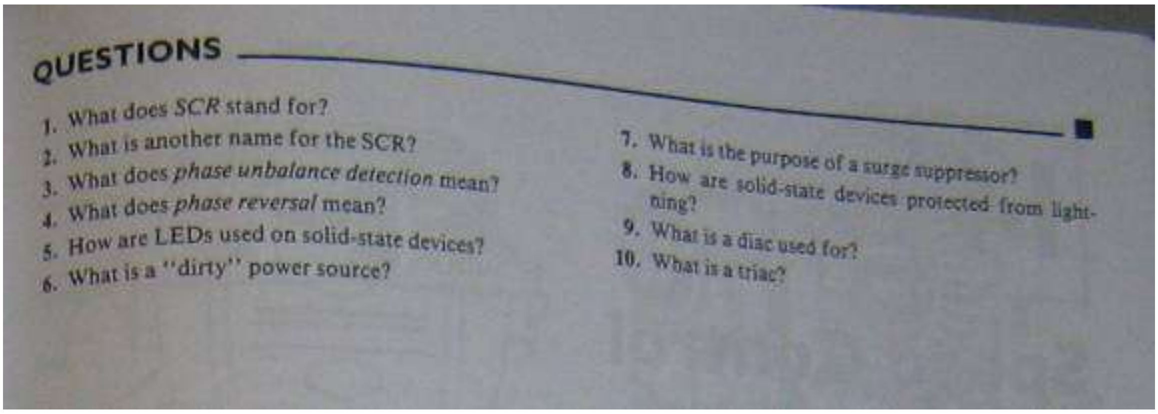
Lightning can also present some high-voltage surges. Secondary surge arrestors can be installed to prevent problems associated with lightning (Fig. 13-28). The arrestor shown in Fig. 13-28A is for single-phase two- or three-wire grounded service. Two of them may be installed to provide protection on 208Y/120V ac three-phase four-wire services. This suppressor will handle 1500 A at 940 V, 5000 A at 1600 V, 10,000 A at 2200 V, and 20,000 A at 3250 V.

A suppressor for use on 650-V ac phase-to-ground maximum is shown in Fig. 13-28B. It is used for three- or four-wire grounded service such as single-phase three-wire, three-phase three-wire, or three-phase four-wire systems. This suppressor will handle 1500 A at 2200 V, 5000 A at 2900 V, and 10,000 A at 3400 V, and 20,000 A at 4000 V. These are maximum discharge voltages that appear across the arrestor during the passage of the discharge current. Discharge current is the current at the arrester during sparkover.

FIGURE 13-28 *(A) Secondary surge arrestors used in lightning protection for electrical systems; 175-V ac phase-to-ground maximum. (B) Secondary surge arrestors for lightning protection used for electrical systems; 650-V ac phase-to-ground maximum. (Square D)*

(A) (B)





QUESTIONS

1. What does *SCR* stand for?
2. What is another name for the SCR?
3. What does *phase unbalance detection* mean?
4. What does *phase reversal* mean?
5. How are LEDs used on solid-state devices?
6. What is a "dirty" power source?
7. What is the purpose of a surge suppressor?
8. How are solid-state devices protected from lightning?
9. What is a diac used for?
10. What is a triac?

Learning Outcome 3.4

Contrast the starting techniques used for a slip ring motor with a squirrel cage motor

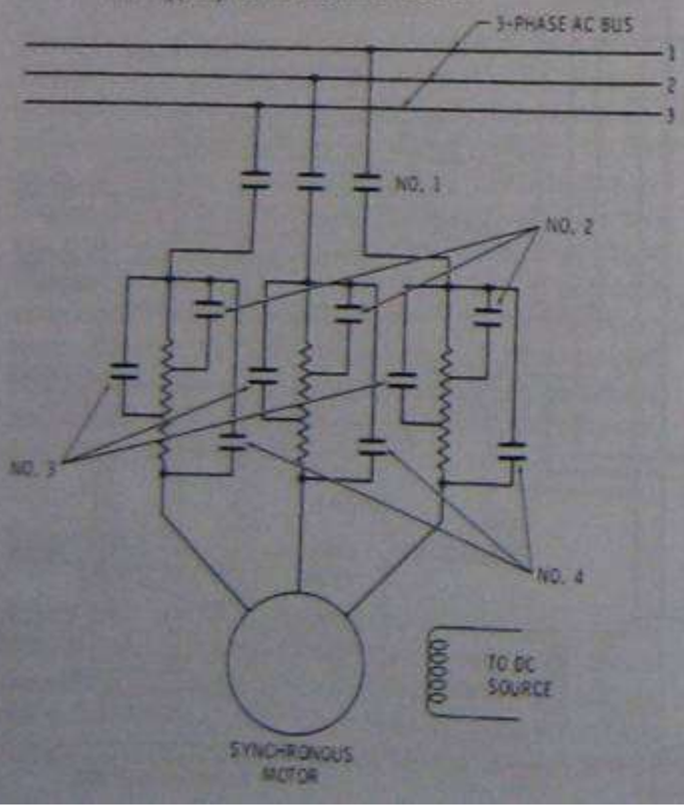
Learning Outcome 3.5

State the applications where a slip ring motor can be used

Reactance Starting. Reactance starting is similar to the reduced-voltage starting methods, except that the first step is obtained by reactance in series with the motor armature instead of autotransformers. In the reactance method of starting, more current is required from the line for the same torque on the first step than when compensators are used. It has one advantage. No circuit opening is required when the motor is transferred to running voltage. The transfer is accomplished by short-circuiting the reactance.

Resistance Starting. A typical circuit using the resistance method of reduced-voltage starting is shown in Fig. 14-6. Switch 1 is closed first. This connects the motor to the line through the entire resistance. Switches 2, 3, and 4 are then closed, with a time interval between each closing. Each switch, in turn, short-circuits a part of the resistance. This method of start-

FIGURE 14-6 *Schematic diagtram of a resistance-type synchronous motor starter.*



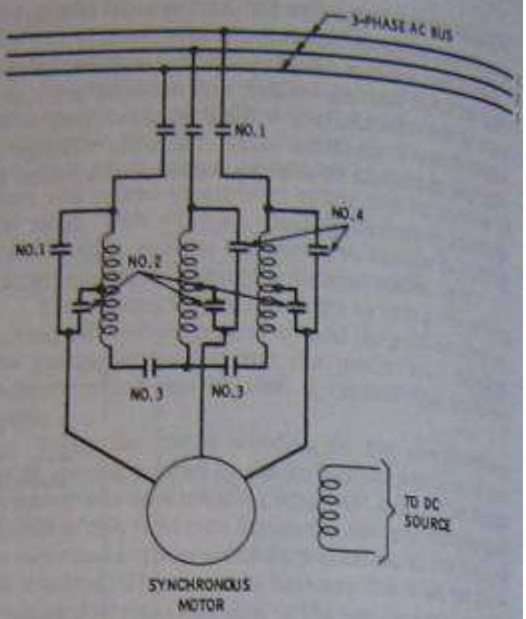


FIGURE 14–7 *Schematic diagram of a Korndorfer-type synchronous motor starter.*

ing is sometimes used when power company rulings require several progressive steps of starting current.

Korndorfer Starting. The reactance and resistance methods are similar to starting the motor using the *Korndorfer method.* It permits the motor to be started without opening the motor circuit. The motor is first connected through suitable taps of a compensator, and then started by connecting the compensator to the line. Full voltage is connected by first opening the neutral of the starting compensator. This allows the motor to run with part of the compensator winding in series with the motor. Then the entire compensator winding is short circuited (Fig. 14–7).

Switch 1 is closed first. This connects one of the compensator windings to the line. Then switch 2 is closed, completing the motor circuit at reduced voltage. As the motor increases its speed, a timing relay, operated by switch 2, opens the circuit of 3. This, in turn, opens the transformer neutral. Switch 4 is closed next. This connects the motor to full-line voltage by shorting the compensator sections. By opening switch 2, the reduced-voltage taps of the compensator are disconnected and the permanent running connection to the motor is completed.

Other Methods of Starting. An auxiliary prime mover, usually an induction motor, may be used as a starter. This method of starting is applied to the motors that have no squirrel-cage winding, or it is used with alternators converted to motor use. This type of motor cannot start under load.

Uses for Synchronous Motors

Synchronous motors may be used for power factor correction; for constant-speed, constant-load drives; and for voltage regulation. Because of the higher efficiency possible with synchronous motors, they can be used advantageously on most loads where constant speed is required. Typical applications are compressors, fans, blowers, line shafts, centrifugal pumps, rubber and paper mills, and to drive dc generators.

WOUND-ROTOR MOTORS

The wound-rotor motor differs from the squirrel-cage type. It has wire-coil windings in its rotor instead of a series of conducting bars in the rotor. Inserting external resistance in the motor circuit when starting will develop a high torque with a comparatively low starting current. As the motor comes up to speed, the resistance is gradually removed until, at full speed, the rotor is short-circuited. Speed can be regulated, within limits, by varying the amount of resistance in the rotor circuit (Fig. 14–8).

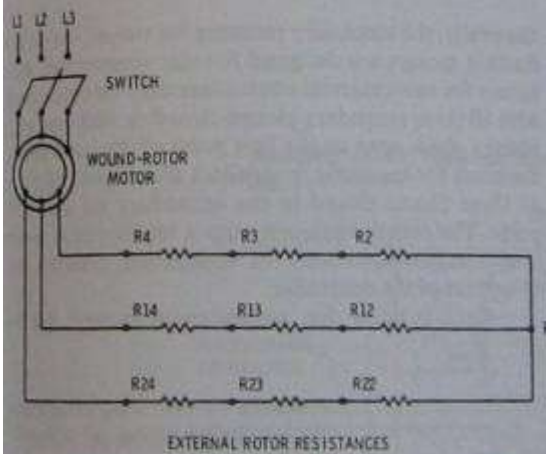


FIGURE 14–8 *Wiring diagram with resistor connections.*

Speed Regulation by Resistance

Resistors can be used to regulate the speed if they are of the proper size to prevent overheating from constant use. The resistors used in starting are used only for a short time, but those used for continuous motor speed reduction are in use for longer periods of time. This means that a resistor must be selected for its intended purpose.

Dc motors produce the most effective variable-speed outputs. However, wound-rotor motors, because of their adjustable rotor resistance, are one of the few means of speed control available for ac motors.

Wound-rotor motors are just that—they have a wound rotor. They are insulated coils of wire that are not permanently short circuited, as in the squirrel-cage motor, but are connected in regular succession to form a definite polar area having the same number of poles as the stator. The ends of these rotor windings are brought out to collector rings, usually referred to as *slip rings*.

Currents induced in the rotor are carried by means of slip rings (and carbon brushes riding on the slip rings) to an externally mounted resistance (Fig. 14–9). These resistances can then be regulated or changed according to the needs of the start sequence. By changing the resistance in the rotor circuit it is possible to change the speed of the motor. Once it has come up to synchronous speed the resistors are then short circuited and the motor runs with characteristics similar to a squirrel-cage type.

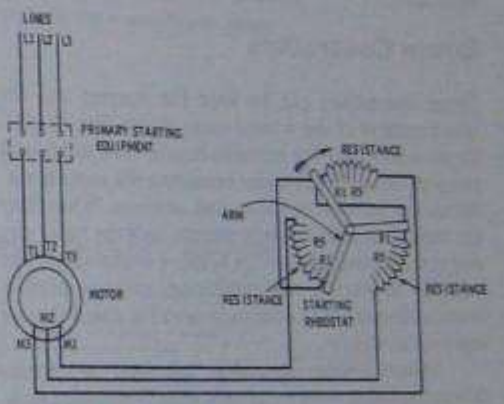


FIGURE 14–9 *Starter–controller for a wound-rotor induction motor.*

However, some resistance can be left in the circuit to aid in speed control, that is, of course, if the size (wattage rating) of the resistors is such as to withstand the constant current flow through them. By placing a high resistance in the rotor circuit, it is possible to start the motor and have it produce high starting torque with low starting current.

Types of Speed Control

The wound-rotor motor can be used where the speed range is small, where the speeds desired do not coincide with a synchronous speed of the line frequency, and where the speed must be gradually or frequently changed from one value to another. This includes compressors, pulverizers, stokers, and conveyors.

A smooth, no-jerk start can be obtained by using the wound-rotor motor. It is simply a matter of supplying the right control equipment.

Multiswitch Starters

Figure 14-10 shows how a typical multiswitch starter for a wound rotor is wired into the rotor circuit. This type of starter is used in the secondary circuits of large wound-rotor induction motors up to 2000 hp with rotor currents up to 1000 A. Contact levers are of the double-pole type and are mechanically arranged in such a manner that they must be closed in a predetermined sequence, and only one at a time. Since the switches are designed for hand-over-hand operation, a desirable time element is introduced that prevents too-rapid acceleration of the motor. When the final switch has been closed, it is held in place by a magnetic coil, and because of the mechanical interlocking feature, all other switches remain closed. This type of starter is just that—a starter, it is not useful as a speed regulator.

Drum Controllers

Drum controllers can be used for starting and for speed control of the wound-rotor motor (Fig. 14-11). Drum controllers are made to handle both stator and rotor circuits. The cylinder mounting the contact segments are built in two insulated sections. When they are built to handle the rotor circuit, only the stator circuit is controlled by a circuit breaker or line starter. In addition to starting and regulating, speed-regulation drum collectors are commonly used for speed-reversing duty as well.

Motor-driven controllers are used in certain drives requiring close automatic speed regulation such as in large air-conditioning plants, blowers, stokers, and similar applications. Some of these installations have been in use for a number of years and are gradually being replaced by more modern motor control methods.

Magnetic Starters

Magnetic starters are built to regulate motor speed, start the motor, and to set the speed of the motor. They consist of a magnetic contactor for connecting the stator circuit to the line, and one or more accelerating contactors to commutate the resistance in the rotor circuit. The number of secondary accelerating contactors varies with the rating, a sufficient number being used to assure smooth acceleration and to keep the inrush current within practical limits. The operation of the accelerating contactors is controlled by a timing device, which provides definite time acceleration. For high-voltage service, the primary contactor is usually of the oil-immersed type. The diagram of a typical magnetic starter for use with a wound-rotor induction motor is shown in Fig. 14-12.

Resistors

Generally, the secondary resistors for wound-rotor induction motors are designed for star connection. Resistors for most manual controllers may be connected with all three secondary phases closed or with one secondary phase open on the first point of the controller. Resistors for magnetic controllers are connected with all three phases closed in the secondary on the first point. The torque obtained with a resistor of a given class number varies with the connection used on the first point of the controller.

Keep in mind that wound-rotor motors can be

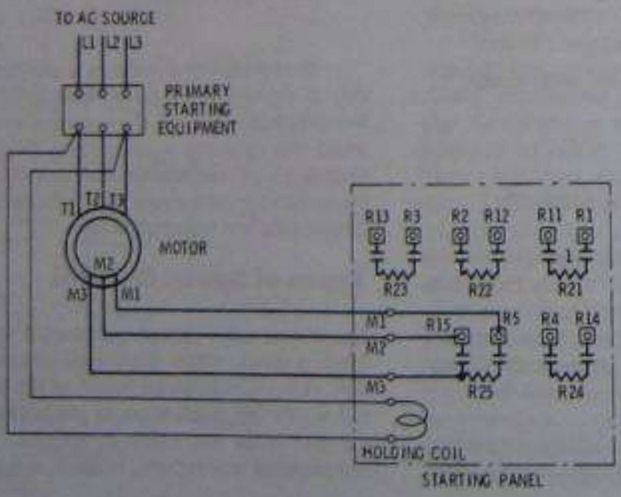


FIGURE 14-10 *Diagram of a typical multiswitch starter for a wound-rotor motor.*

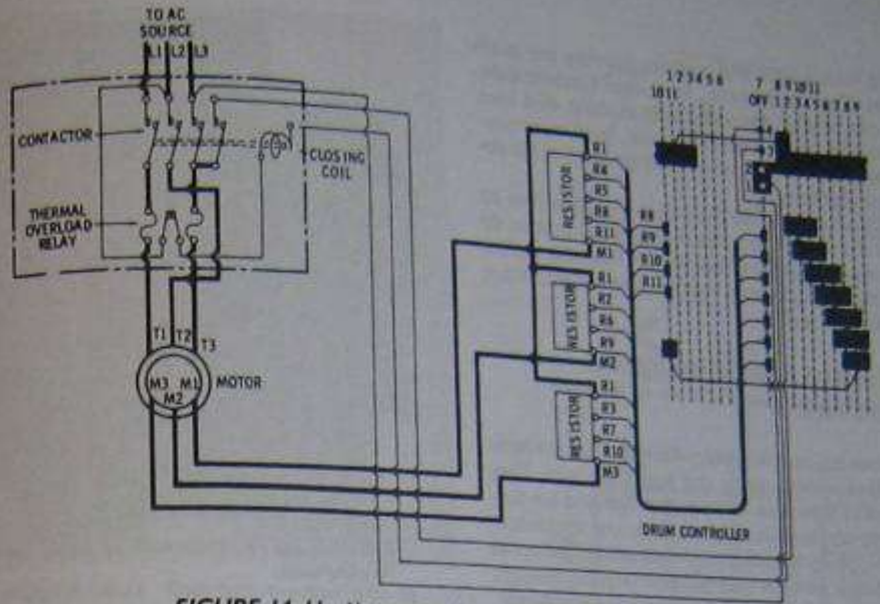
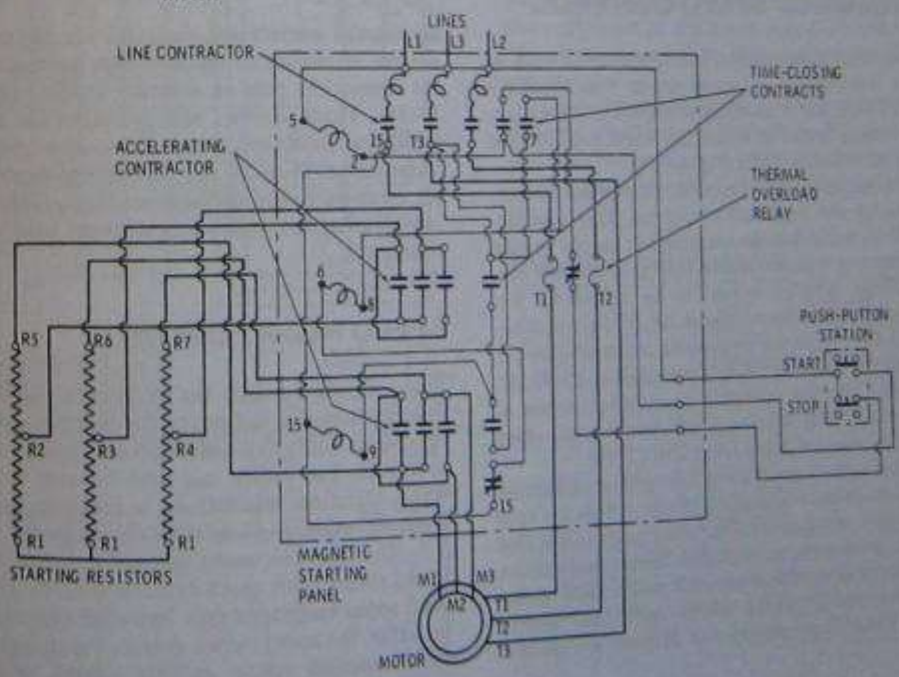


FIGURE 14–11 Nonreversing drum controller for a wound-rotor motor with a three-phase secondary.

FIGURE 14–12 Magnetic starter contactor for use with a wound-rotor motor.



Learning Outcome 3.4

Contrast the starting techniques used for a slip ring motor with squirrel cage motor

Learning Outcome 3.6

List the advantages and disadvantages of two types of motors

Learning Outcome 3.7

Draw the schematic diagrams for the various types of induction motor braking circuits

17.14 The synchronous motor versus the induction motor

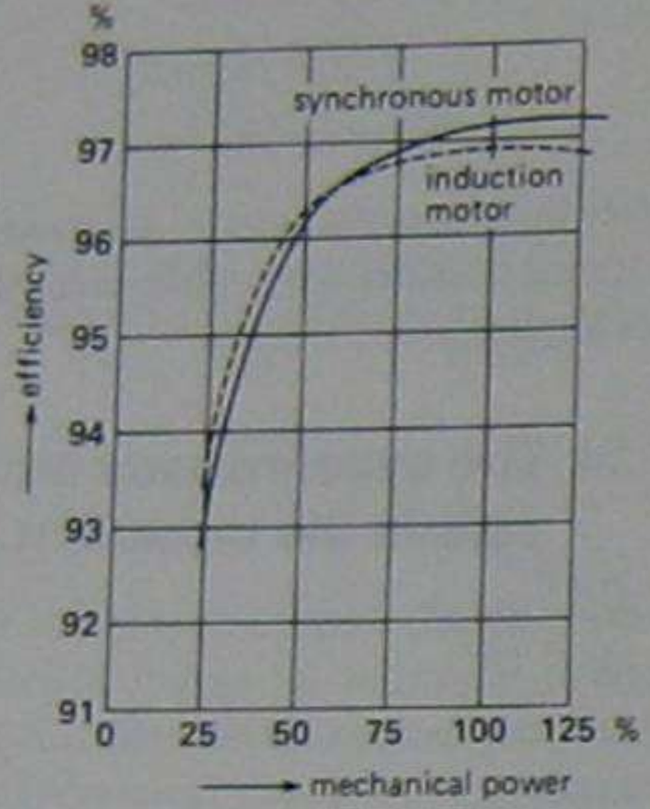
We have already seen that induction motors have excellent properties for speeds above 600 r/min. But at lower speeds they become heavy, costly, and have relatively low power factors and efficiencies.

Synchronous motors are particularly attractive for low-speed drives because the power factor can always be adjusted to 1.0 and the efficiency is high. Although more complex to build, their weight and cost are often less than those of induction motors of equal power and speed. This is particularly true for speeds below 300 r/min.

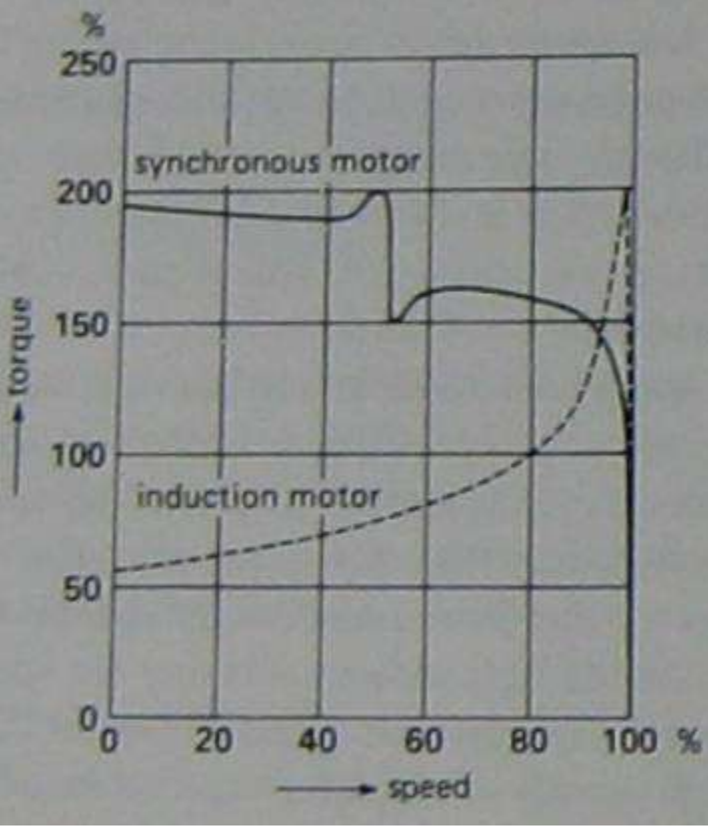
A synchronous motor can improve the power factor of a plant while carrying its rated load. Furthermore, its starting torque can be made considerably greater than that of an induction motor. The reason is that the resistance of the squirrel-cage winding can be high without affecting the speed or efficiency at synchronous speed. Figure 17.23 compares the properties of a squirrel-cage induction motor and a synchronous motor having the same nominal rating. The biggest difference is in the starting torque.

High-power electronic converters generating very low frequencies enable us to run synchronous motors at ultra-low speeds. Thus, huge motors in the 10 MW range drive crushers, rotary kilns, and variable-speed ball mills.

(a)


%
98
97
96
95
94
93
92
91
efficiency
synchronous motor
induction
motor
0
25
50
75
100
125 %
mechanical power

(b)


%
250
200
150
100
50
0
torque
synchronous motor
induction motor
0
20
40
60
80
100 %
speed

When Starting Any One Requires Another

Several motors can be run independently of each other with some of the starters actuated by two-wire and some by three-wire pilot devices. Whenever any one of these motors is running, a pump or fan motor must also run (Fig. 15–9).

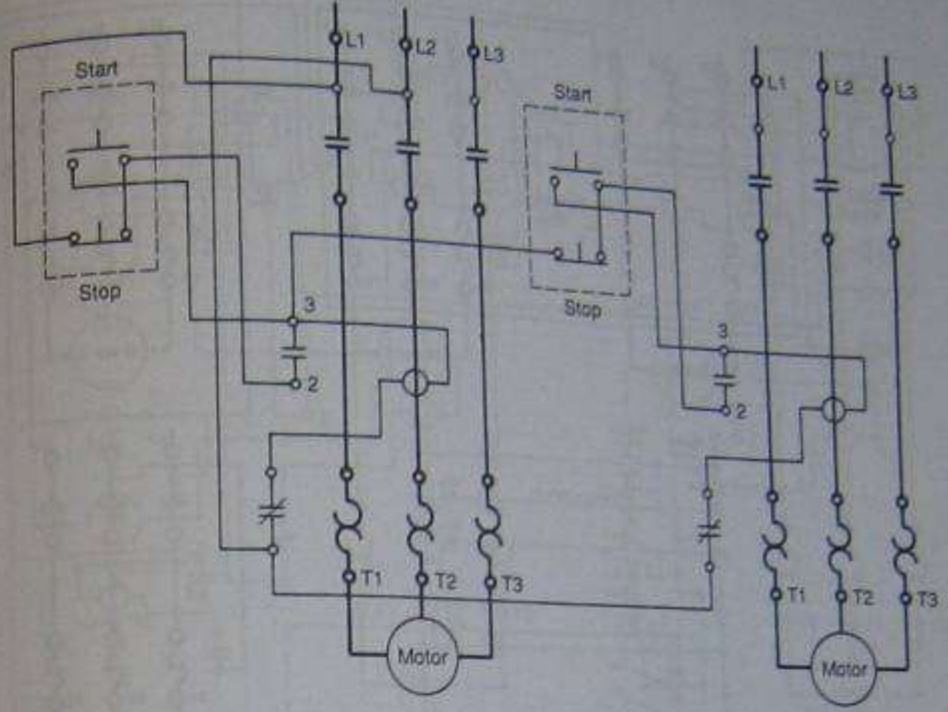
A master start–stop pushbutton station with a control relay is used to shut down the entire system in an emergency. Control relay (CR) provides three-wire control for M1, which is controlled by a two-wire control device such as a pressure switch. Motors M2 and M3 are controlled by start–stop pushbutton stations.

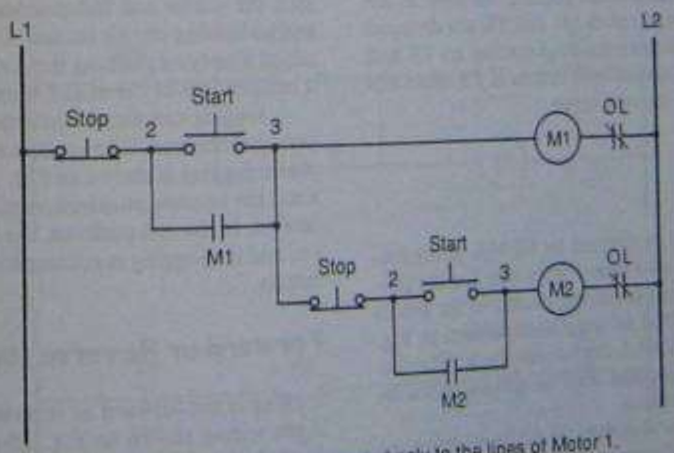
Auxiliary contacts on M1, M2, and M3 control M4. These auxiliary contacts are all wired in parallel so that any one of them may start M4. On some starters auxiliary contacts have been added to M2 and M3 for this purpose. The standard *hold-in contact* on M1 may be used as an auxiliary if wire Y is removed. Hold-in contacts are not required when a two-wire control device is used.

When this system is used, the phase connections on all of the starters must be the same. That is, L1 of each starter must be connected to the same incoming phase line; L2 and L3 of each starter must be phased out similarly.

Automatic Sequence Control

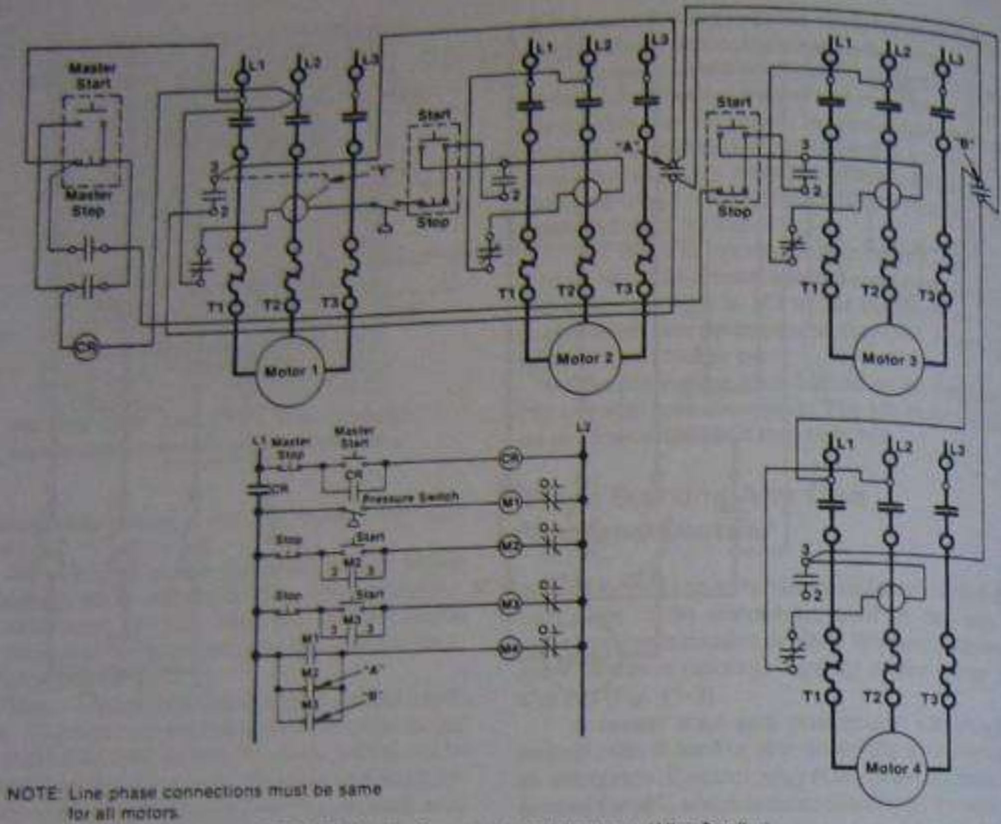
Having automatic sequence control is also possible with the arrangement shown in Fig. 15–10. In this system it is desired to have a second motor start automatically when the first one is stopped. The second motor is to run only for a given length of time. A good application of this might be found where the second motor is needed to run a cooling fan or a pump after the first





NOTE: Control circuit is connected only to the lines of Motor 1.

FIGURE 15–8 *Sequence control diagrams. (Allen-Bradley)*



NOTE: Line phase connections must be same for all motors.

FIGURE 15–9 *Sequence control diagrams. (Allen-Bradley)*

To accomplish this, an off-delay timer (TR) is used. When the START button is pressed, it energizes both M1 and TR. This operation of TR closes its time-delay contact, but the circuit to M2 is kept open by the opening of the instantaneous contact. As soon as the STOP button is pressed, both M1 and TR are dropped out. This closes the instantaneous contact on TR and starts M2. M2 will continue to run until TR times out and the time-delay contact opens.

JOGGING

Jogging, or inching, is defined by NEMA as the momentary operation of a motor from rest for the purpose of accomplishing small movements of the driven machine. One method of jogging is shown in Fig. 15-11. The selector switch disconnects the holding circuit interlock and jogging may be accomplished by pressing the START button.

There are several means of accomplishing the jogging operation. Figure 15-12 shows how jogging is done using a control relay. Pressing the START button energizes the control relay that in turn energizes the starter coil. The normally open starter interlock and relay contact then form a holding circuit around the START button. However, pressing the JOG button energizes the starter coil independent of the control relay and no holding circuit forms. Then jogging can be obtained simply by pushing the JOG button and releasing it independent of the START button.

Jogging can also be accomplished by using a *selector* pushbutton. The use of a selector pushbutton to obtain jogging is shown in Fig. 15-13. In the RUN position the selector pushbutton gives normal three-wire control. In the JOG position, the holding circuit is broken and the jogging is accomplished by depressing the button.

Forward or Reverse Jogging

Jogging in the forward or reverse direction is possible if the wiring shown in Fig. 15-14 is followed. This control scheme permits jogging the motor either in the forward or reverse direction, whether the motor is at a standstill or is rotating in either direction. Pressing the

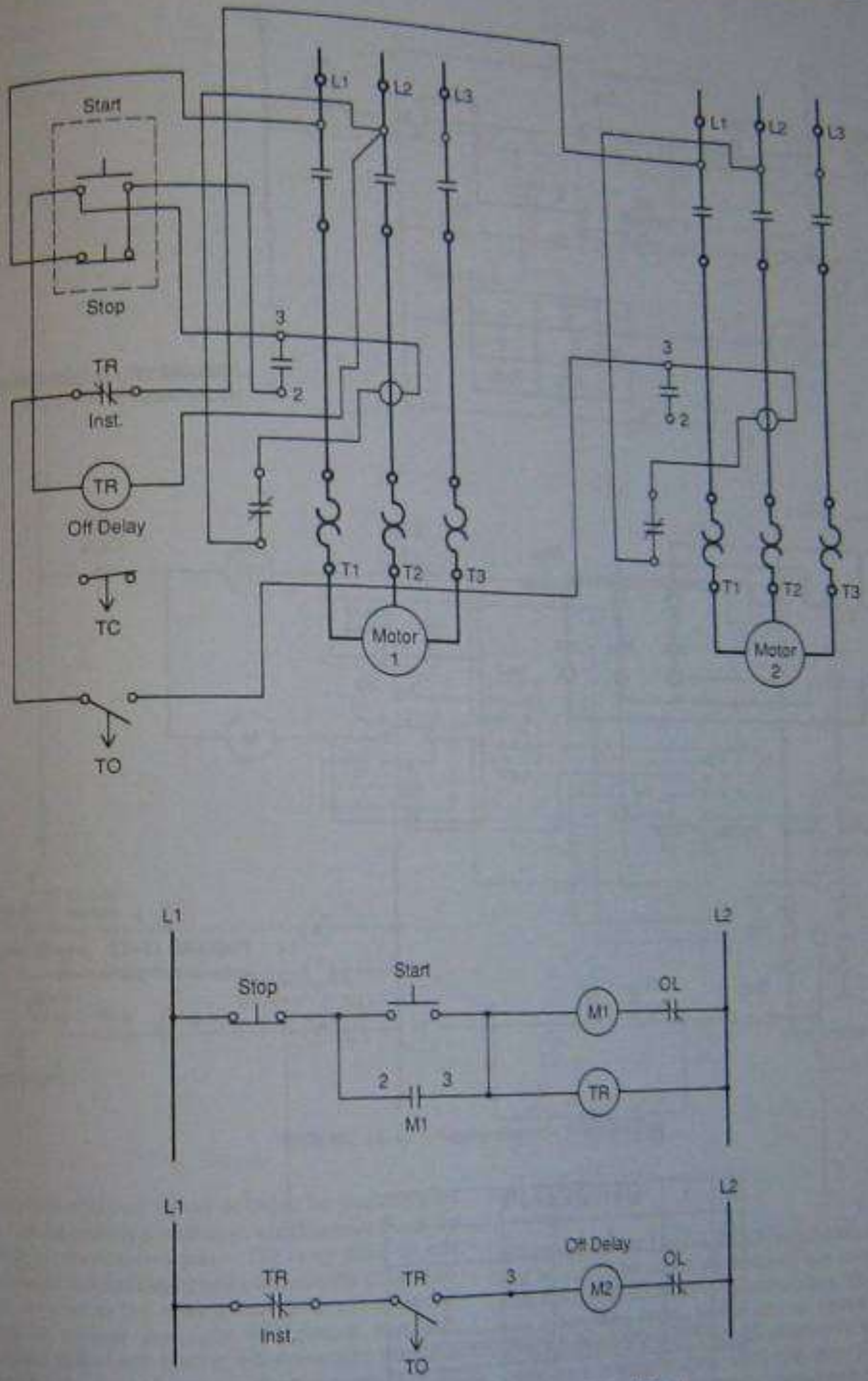
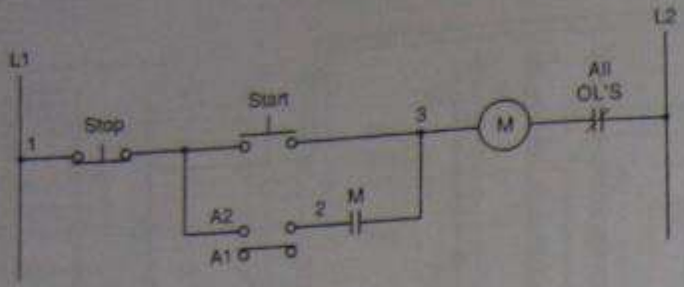


FIGURE 15–10 *Sequence control diagrams. (Allen-Bradley)*



A1	1	
A2		1
	JOG	RUN

FIGURE 15–11 Jogging with a selector switch.

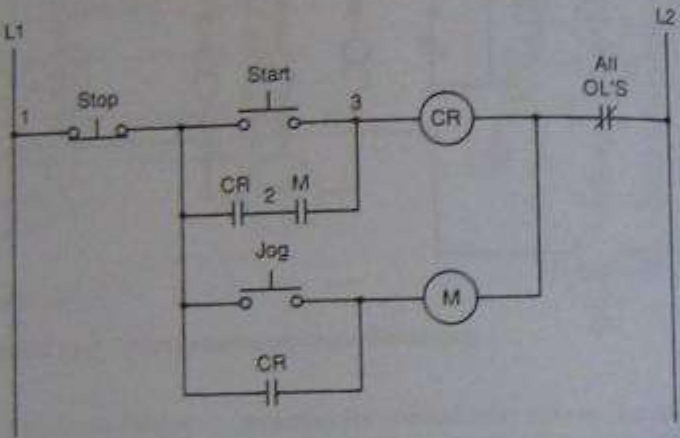


FIGURE 15–12 Jogging with a control relay.

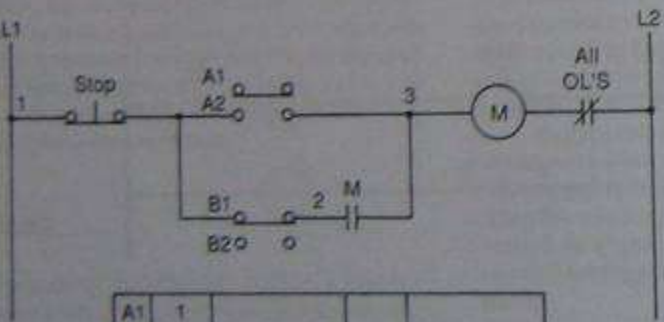


FIGURE 15–13 Jogging using a selector switch pushbutton.

A1	1			
A2		1		1
B1	1	1		
B2				1
	FREE	DEPRESSED	FREE	DEPRESSED
	RUN		JOG	

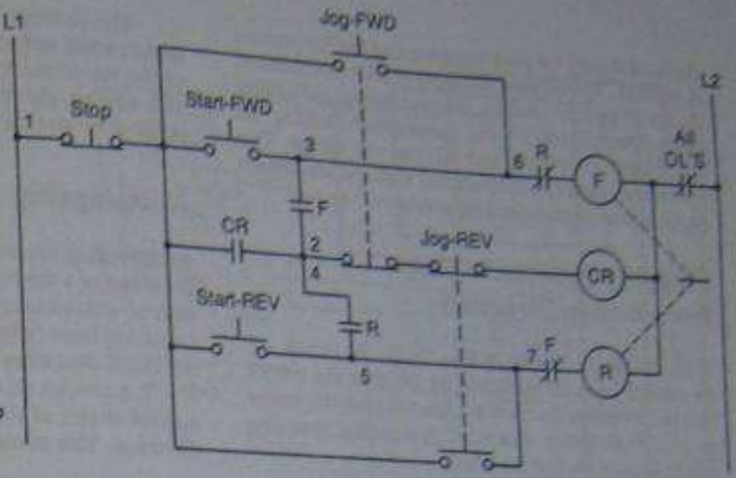


FIGURE 15-14 Jogging using a control relay for reversing starter.

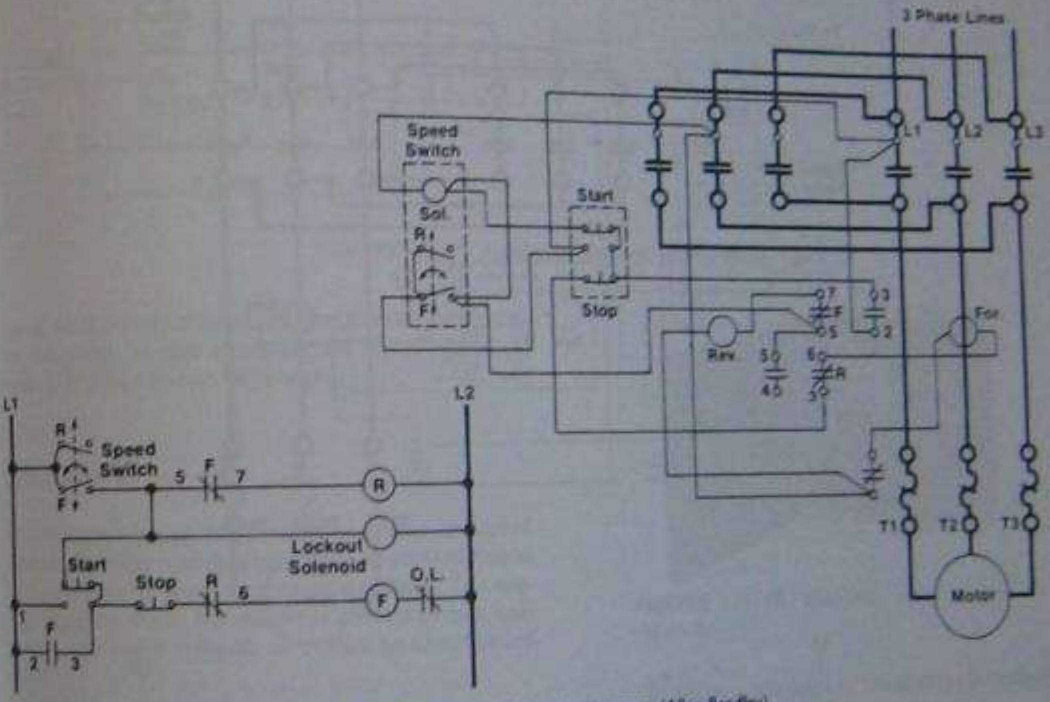


FIGURE 15-15 Plugging diagrams. (Allen-Bradley)

START-FORWARD or START-REVERSE buttons energizes the corresponding starter coil, which in turn closes the circuit to the control relay. The relay picks up and completes the holding circuit around the START button. As long as the relay is energized, either the forward or reverse contactor will remain energized. Pressing *either* JOG button will deenergize the relay, releasing the closed contactor. Further pressing of the JOG button permits jogging in the desired direction.

PLUGGING

Plugging is defined by the NEMA as a system of *braking* in which the motor connections are reversed so that the motor develops a countertorque. Thus it exerts a retarding force. In the scheme shown in Fig. 15-15 the motor is run in one direction only and must come to a complete stop when the STOP button is pressed. The reverse contactor of the reversing switching is used only for plug stopping and not for running in reverse. The lockout solenoid is built into some of the speed switches and its function is to guard against an accidental turn of the motor shaft, closing the speed switch contacts and starting the motor. This protective feature is optional and the speed switch can be furnished without lockout solenoid if desired.

Plugging a Motor to Stop from Either Direction

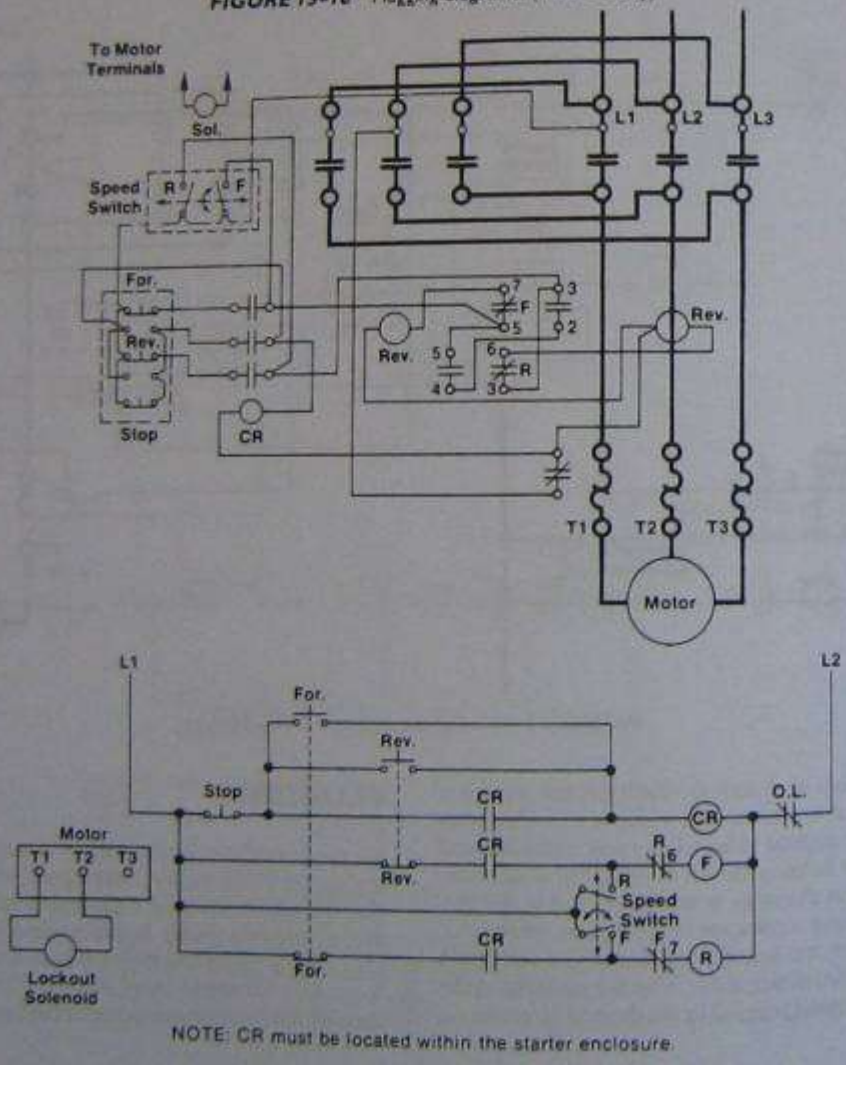
With the system shown in Fig. 15-16, the motor can be started in either direction by pressing the proper button. Pressing the STOP button will plug the motor to stop from either direction. A standard reversing switch is used for this purpose.

The lockout solenoid is a built-in part of the speed switch and it guards against an accidental turn of the motor shaft closing the speed switch contacts and starting the motor. The control relay and the pushbutton station are standard parts.

Antiplugging

Antiplugging protection is defined by the NEMA as the effect of a device that operates to prevent application of countertorque by the motor until the motor speed has been reduced to an acceptable value. With the motor operating in one direction, as shown in Fig. 15-17, a contact on the antiplugging switch opens the control circuit of the contactor used for the opposite direction. This contact will not close until the motor

FIGURE 15-16 Plugging diagrams. (Allen-Bradley)



NOTE: CR must be located within the starter enclosure.

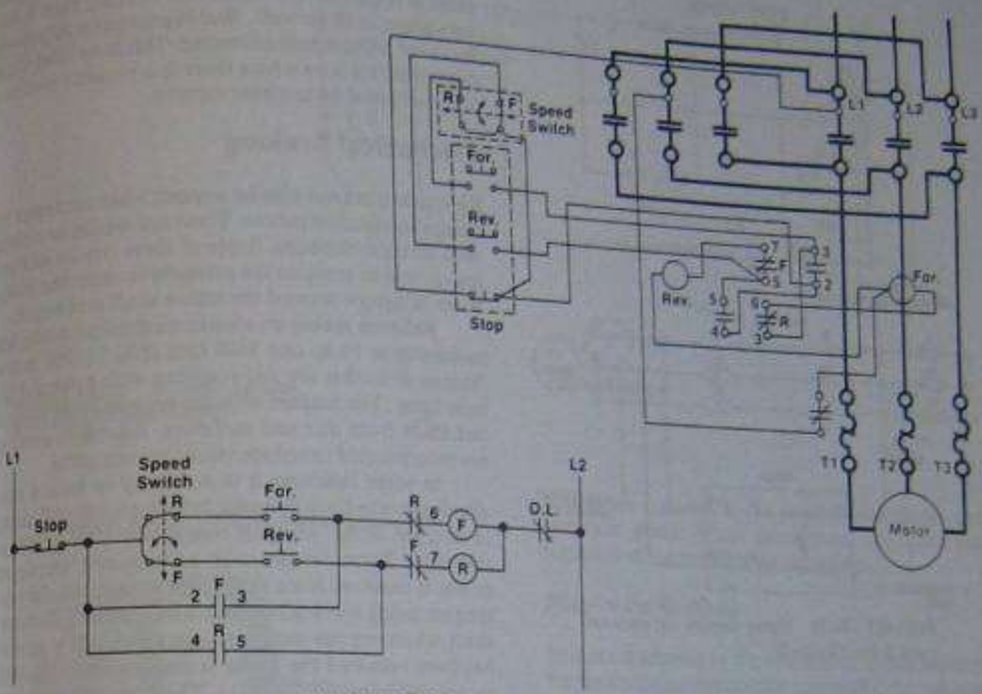


FIGURE 15-17 Antiplugging diagrams. (Allen-Bradley)

has slowed down, after which the other contactor can be energized. In this schematic the motor can be reversed, but it must not be plugged.

BRAKING

Electric motors can be brought to a stop or braked both electrically and mechanically. In some instances it is necessary to use a combination of both. This usually happens when the motor is connected to a load that is not easily stopped or cannot be disconnected easily.

Electronic Motor Brake

The electronic motor brake made by Square D provides a simple, effective means of braking an ac squirrel-cage motor (Fig. 15-18). It can be used for woodworking machines such as saws and sanders, and for machine tools such as lathes and drills, as well as for conveyor systems, textile machinery, and centrifuges. Heating, venting, and air-conditioning fans and many other machines in varied industries may also use this type of braking.

FIGURE 15-18 Electronic motor brake. (Square D)

The major advantages of the electronic methods versus the mechanical brake system are:

1. No friction, wear, or maintenance
2. Adjustable soft-stop capability
3. No mechanical connection to the motor shaft
4. Multimotor braking capability
5. Easily wired to a new or existing machinery
6. Unaffected by hostile motor environment

Electronic braking is commonly known as dynamic braking. Dynamic braking of an ac induction

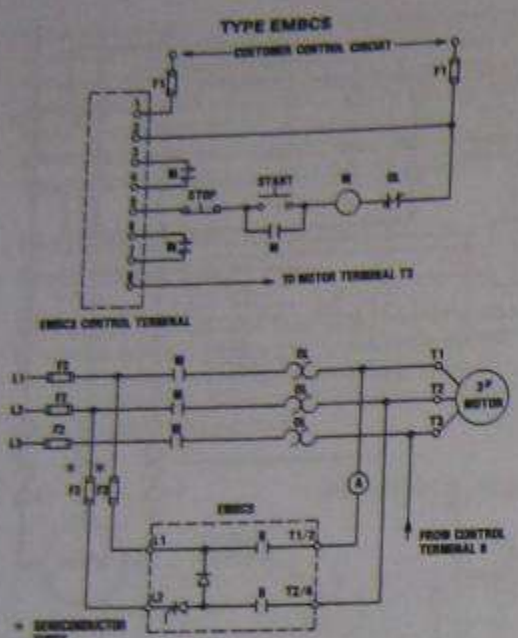


FIGURE 15-19 Wiring diagram for electronic motor brake. (Square D)

motor is generally accomplished by exciting its stator windings with dc current. The amount of braking torque is directly proportional to the dc current passing through the stator windings of the motor (Fig. 15-19).

Dynamic braking of a motor may cause threaded fasteners connected to the motor shaft to loosen, due to the reverse torque applied. Use positive-locking fasteners or fastening compound to prevent such loosening.

Note: Electronic motor brakes will not stop the motor if power is lost or disconnected.

This type of electronic motor brake can be used to stop a load and signal a mechanical brake system to hold it. In addition, the brake will interface with either jogging, reversing, multispeed, or reduced-voltage motor starter applications.

The electronic motor brake is designed such that the braking contactorcloses before the thyristor (SCR) switches the braking current on. The contactor will not open until after the braking current has been switched off. This allows the braking contactor to be rated for current-carrying capacity only and not for the higher make-and-break duty.

An additional circuit detects when the motor has come to a halt, switches off the braking current, and permits the motor to restart. No braking time adjustment is required. The maximum braking time is factory preset at 10 seconds. Braking torque is adjustable by use of a single potentiometer. This is an ideal braking system for jobs where there is a variable load and for multispeed three-phase motors.

Mechanical Braking

Electric motors can also be stopped when necessary by using a mechanical means. These are similar to what is used with automobiles. Some of them rely on an electric current to energize the solenoid to cause the brake shoes to tighten around the motor shaft and stop it.

Reliance makes an electromechanical brake for motors up to 10 hp and 3600 rpm (Fig. 15-20). It has friction disks that are self-resetting with a manual release lever. The magnet coils are encapsulated to protect them from dirt and moisture. Antirattle springs are incorporated to reduce vibration and noise.

In some instances it is necessary to have a mechanical brake since dynamic braking is not sufficient to stop the motor rotation completely after power is removed. These brakes may be actuated whenever power is removed from the motor circuit. An electromagnet holds the brake shoes away from the motor shaft whenever the motor is energized. Once power has been removed the brake is automatically applied by spring action (Fig. 15-21). This type of braking is very useful in elevators and similar installations.

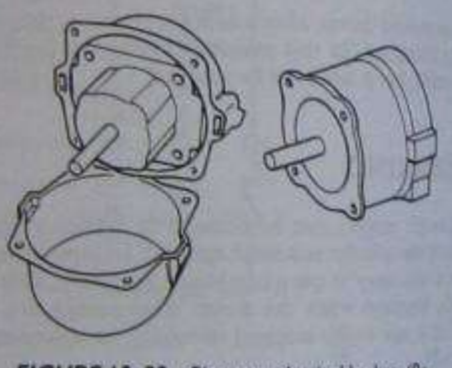


FIGURE 15-20 Electromechanical brakes. (Reliance)

Thruster Brakes

Thruster brakes are used with ac or dc motors and provide a smoothly applied fixed torque for hold or for stopping (Fig. 15-22). They are used on crane travel drives, lift bridges, conveyors, and similar applications to reduce load sway, and affect loading to motors and the mechanical system. These brakes are released by a thruster mechanism. This self-contained

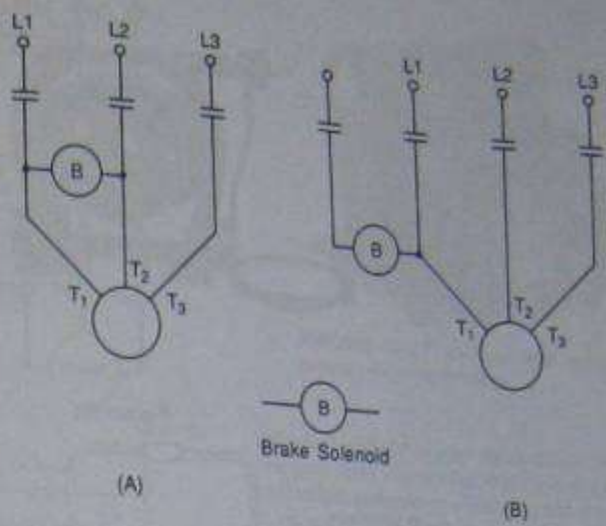
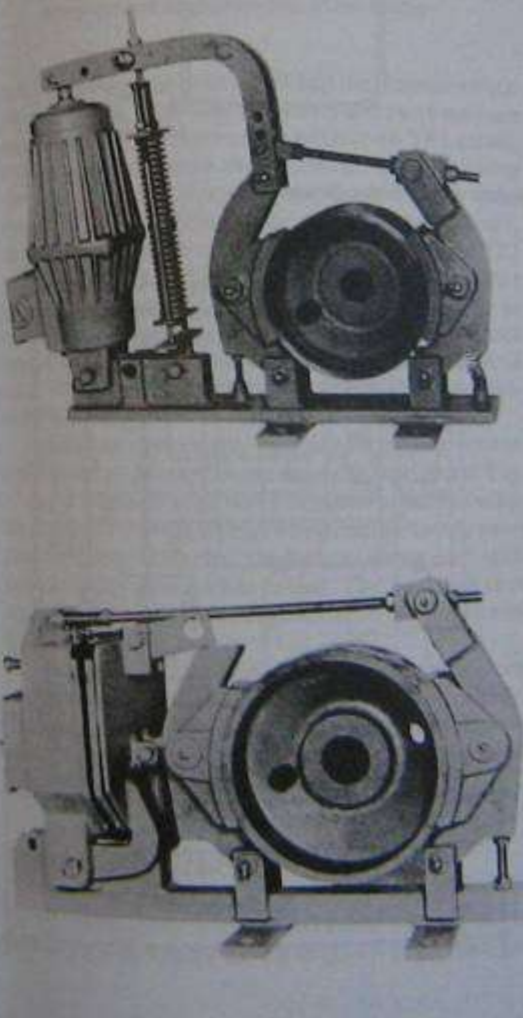


FIGURE 15-21 Ac brake coil hook ups for across-the-line starting.

FIGURE 15-22 Thruster brakes. (Square D)



mechanism contains an ac squirrel-cage motor and hydraulic pump. When deenergized the brake sets smoothly as the pumping action ceases.

Magnetic Brakes

Brakes are selected by the amount of torque required for the particular application. Generally, the full-load torque of the motor is used as a basis for determining the brake torque required. This can be calculated by using the following formula for both ac and dc motors:

$$\text{torque} = \frac{\text{rated hp} \times 5252}{\text{rated rpm}}$$

Depending on the characteristics of the drive, the braking torque required may be more or less than the full-load torque of the motor. In addition to being selected to meet the torque requirements of the particular application, the magnetic brake used for stopping must be selected to prevent overheating of the brake wheel when operated on the anticipated duty cycle.

Hydraulic Brakes

Hydraulic brakes are used with ac or dc motors to provide an operator-controlled infinitely adjustable torque for slowing and stopping. These are used on crane travel drives, mill machines, conveyors, and similar jobs. They are spring released, hydraulically applied shoe-type friction brakes designed to meet AISE (American Iron and Steel Engineers) standards for mounting. The standard brake includes corrosion-resistant hardware and grease fittings (Fig. 15-23). Figure 15-24 shows the typical piping diagram for one brake.

Learning Outcome 4.1

Perform measurements to obtain the synchronous impedance of a synchronous motor

(Background theory)

12.6 Equivalent Circuit of the Synchronous Machine

The preceding section has shown that the equivalent circuit of a synchronous machine must contain a voltage source E_F which is constant for a constant excitation current I_F and a series-connected reactance X_A. In addition, an actual machine winding will have resistance R and (in the same way as a transformer) leakage reactance X_L.

Fig. 12.7(*a*) shows the full equivalent circuit of the synchronous machine in which the current flows in the conventionally positive direction for generator-mode operation (a source), i.e. emerging from the positive terminal. Applying Kirchhoff's law to this circuit,

$$E_F = V + IR + jIX_L + jIX_A \tag{12.19}$$

Fig. 12.7(*b*) is the corresponding complexor diagram. The resultant e.m.f. E_R is shown for the sake of completeness but will be omitted in subsequent diagrams.

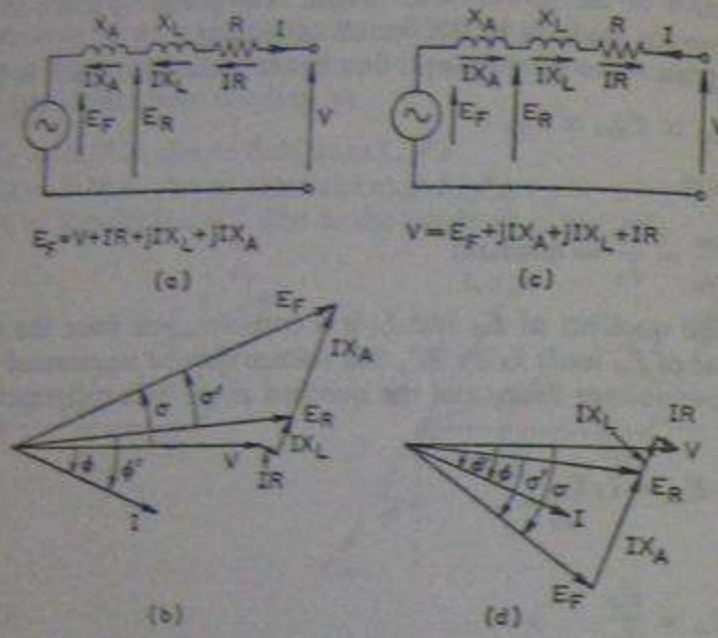


Fig. 12.7 EQUIVALENT CIRCUITS AND FULL COMPLEXOR DIAGRAMS FOR THE SYNCHRONOUS MACHINE
(*a*), (*b*) Generator (*c*), (*d*) Motor

Eqn. (12.19) may be rewritten as

$$E_F = V + IZ_s \tag{12.20}$$

where Z_s is the *synchronous impedance*.

$$Z_s = R + j(X_L + X_A) \tag{12.21}$$

or

$$Z_s = R + jX_s \tag{12.22}$$

where X_s is the *synchronous reactance*:

$$X_s = X_L + X_A \tag{12.23}$$

In polar form the synchronous impedance is

$$Z_s = Z_s\underline{/\psi} \tag{12.24}$$

where

$$\psi = \tan^{-1}\frac{X_s}{R} \tag{12.25}$$

and

$$Z_s = \sqrt{(R^2 + X_s^2)} \tag{12.26}$$

armature m.m.f. per pole. This fixed value of armature m.m.f. has a progressively smaller effect as the air-gap is lengthened.

EXAMPLE 12.2 A 3-phase 13·8 kV 100 MVA 50 Hz 2-pole star-connected cylindrical-rotor synchronous generator has an internal stator diameter of 1·08 m and an effective core length of 4·6 m. The machine has a synchronous reactance of 2 p.u. and a leakage reactance of 0·16 p.u. The average flux density over the pole area is approximately 0·6 Wb/m². Estimate the gap length.

Assume that the radial air-gap is constant and the armature winding uniform. Neglect the reluctance of the iron core and the space harmonics in the armature m.m.f.

With the above assumptions the reactance X_A is

$$X_A = \omega\left(\frac{18}{\pi^2}\right)^2 \frac{\mu_0}{3l_g}\pi DL\left(\frac{N_p}{2p}\right)^2 \qquad (12.31)$$

Base voltage, $V_B = V_p = \dfrac{13{\cdot}8 \times 10^3}{\sqrt{3}} = 7{,}960\,\text{V}$

Base current, $I_B = \dfrac{\text{VA/phase}}{V_B} = \dfrac{100 \times 10^6}{3 \times 7{,}960} = 4{,}180\,\text{A}$

Base impedance, $Z_B = \dfrac{V_B}{I_B} = \dfrac{7{,}960}{4{,}180} = 1{\cdot}91\,\Omega$

$X_{Apu} = X_{spu} - X_{Lpu} = 2{\cdot}00 - 0{\cdot}16 = 1{\cdot}84\,\text{p.u.}$

$X_A = X_{Apu}Z_B = 1{\cdot}84 \times 1{\cdot}91 = 3{\cdot}52\,\Omega$

Flux per pole, $B_{av} \times \text{Pole area} = B_{av}\dfrac{\pi DL}{2} = \dfrac{0{\cdot}6 \times \pi \times 1{\cdot}08 \times 4{\cdot}6}{2}$

$= 4{\cdot}68\,\text{Wb}$

$$E_p = K_d K_s \frac{\omega \Phi N_p}{\sqrt{2}} \qquad (11.20)$$

For a uniform winding, $K_d = 3/\pi$ and $K_s = 1$, so that

$$N_p = \frac{\sqrt{2}E_p}{K_d K_s \omega \Phi} = \frac{\sqrt{2} \times 7{,}960}{3/\pi \times 2\pi \times 50 \times 4{\cdot}68} = 8{\cdot}02$$

The number of turns per phase must be an integer, say 8. This will require a slightly higher flux per pole and average value of flux density. From eqn. (12.31),

$$l_g = 2\pi \times 50 \times \left(\frac{18}{\pi^2}\right)^2 \times \frac{4\pi \times 10^{-7}}{3 \times 3{\cdot}52} \times \pi \times 1{\cdot}08 \times 4{\cdot}6 \times \left(\frac{8}{2}\right)^2$$

$$= \underline{3{\cdot}10 \times 10^{-2}\,\text{m}}$$

12.8 Determination of Synchronous Impedance

The ohmic value of the synchronous impedance, at a given value of excitation may be determined by open-circuit and short-circuit tests (Fig. 12.8).

On open-circuit the terminal voltage depends on the field excitation and the magnetic characteristics of the machine. Fig. 12.9 includes a

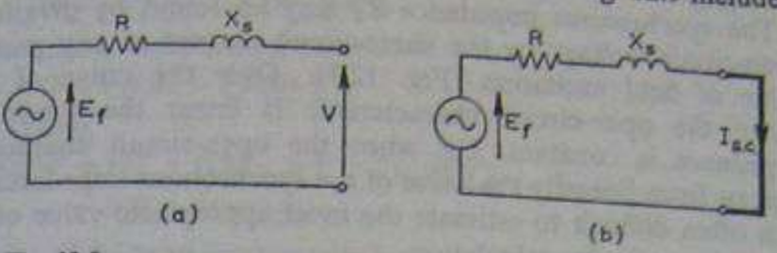


Fig. 12.8 DETERMINATION OF SYCHRONOUS IMPEDANCE
(*a*) Open-circuit test (*b*) Short-circuit test

typical open-circuit characteristic showing the usual initial linear portion and subsequent saturation portion of a magnetization curve.

On short-circuit the current in an alternator winding will normally lag behind the induced voltage by approximately 90° since the leakage

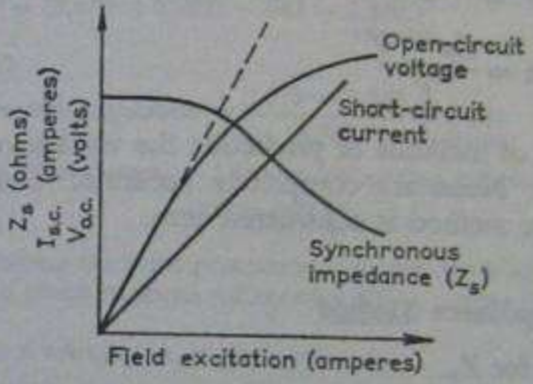


Fig. 12.9 VARIATION OF SYNCHRONOUS IMPEDANCE WITH EXCITATION

reactance of the winding is normally much greater than the winding resistance. The complexor diagram for short-circuit conditions is shown in Fig. 12.10. It is found that the armature and field m.m.f.s

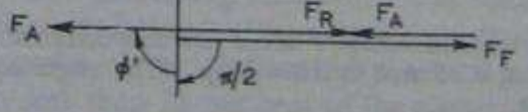


Fig. 12.10 COMPLEXOR DIAGRAM FOR SHORT-CIRCUIT CONDITIONS

are directly in opposition, so that a surprisingly large excitation is required to give full-load short-circuit current in the windings. The resultant m.m.f. and flux are small since the induced voltage is only required to overcome resistance and leakage reactance voltage drops in the windings. Since the flux is small, saturation effects will be negligible and the short-circuit characteristic is almost straight.

The synchronous impedance Z_s may be found by dividing the open-circuit voltage by the short-circuit current at any particular value of field excitation (Fig. 12.9). Over the range of values where the open-circuit characteristic is linear the synchronous impedance is constant, but when the open-circuit characteristic departs from linearity the value of the synchronous impedance falls. It is often difficult to estimate the most appropriate value of Z_s to use for a particular calculation.

12.9 Voltage Regulation

The voltage regulation of an alternator is normally defined as the rise in terminal voltage when a given load is thrown off. Thus, if E_P is the induced voltage on open-circuit and V is the terminal voltage at a given load, the voltage regulation is given by

$$\text{Per-unit regulation} = \frac{E_P - V}{V} \tag{12.32}$$

There are a number of methods of predicting the voltage regulation of an alternator. None are completely accurate. Only the synchronous impedance method is considered here.

12.10 Synchronous Impedance Method

Using a suitable value for Z_s,

$$E_P = V + IZ_s \tag{12.20}$$

EXAMPLE 12.3 A 3-phase star-connected alternator has a resistance of 0·5Ω and a synchronous reactance of 5Ω per phase. It is excited to give 6,600V (line) on open circuit. Determine the terminal voltage and per-unit voltage regulation on full-load current of 130A when the load power factor is (*a*) 0·8 lagging, (*b*) 0·6 leading.

It is best to take the phase terminal voltage V as the reference complexor since the phase angle of the current is referred to this voltage. (The magnitude of V is, however, not known): i.e.

$$\text{Phase terminal voltage, } V = V\underline{/0^\circ}$$

The magnitude of the e.m.f E_P is known but not its phase with respect to V; i.e.

$$E_P = E_P\underline{/\sigma^\circ} = \frac{6{,}600}{\sqrt{3}}\underline{/\sigma^\circ} = 3{,}810\underline{/\sigma^\circ}$$

where σ° is the phase of E_P with respect to V as reference.

(*a*) The phase current I lags behind V by a phase angle corresponding to a power factor of 0·8 lagging, i.e.

$$I = 130\underline{/-\cos^{-1} 0{\cdot}8} = 130\underline{/-36{\cdot}9^\circ}\ \text{A}$$

The synchronous impedance per phase is

$$Z_s = (0{\cdot}5 + j5)\,\Omega = 5{\cdot}02\underline{/84{\cdot}3^\circ}\,\Omega$$

In eqn. (12.20),

$$3{,}810\underline{/\sigma^\circ} = V\underline{/0^\circ} + (130\underline{/-36{\cdot}9^\circ} \times 5{\cdot}02\underline{/84{\cdot}3^\circ})$$
$$= V\underline{/0^\circ} + 653\underline{/47{\cdot}4^\circ}$$

Expressing all the terms in rectangular form,

$$3{,}810 \cos\sigma + j\,3{,}810 \sin\sigma = V + j0 + 442 + j482$$

Equating quadrate parts,

$$3{,}810 \sin\sigma = 482$$

whence $\sin\sigma = 0{\cdot}127$ and $\cos\sigma = 0{\cdot}992$

Equating reference parts,

$$3{,}810 \cos\sigma = V + 442$$
$$V = (3{,}810 \times 0{\cdot}992) - 442 = \underline{3{,}340\ \text{V}}$$

and

$$\text{Per-unit regulation} = \frac{3{,}810 - 3{,}340}{3{,}340} = \underline{0{\cdot}141}$$

(*b*) Phase current = 130 A at 0·6 leading with respect to V

$$= 130\underline{/+53{\cdot}1^\circ}$$

Following the same procedure as in part (*a*) it will be found that there is an on-load phase terminal voltage of $\underline{4{,}260\ \text{V}}$. Hence the per-unit regulation, since there is a voltage rise, is given by

$$\frac{3810 - 4260}{4{,}260} = \underline{-0{\cdot}106\,\text{p.u.}}$$

Learn Outcome 4.3

Contrast the difference in performance between synchronous motor fitted with cylindrical rotors and salient pole rotors

12.15 Power/Angle Characteristic of a Synchronous Machine

Fig. 12.15(*a*) is part of the general load diagram for a synchronous machine and shows the complexor diagram corresponding to generation into infinite busbars at a lagging power factor. Fig. 12.15(*b*) is

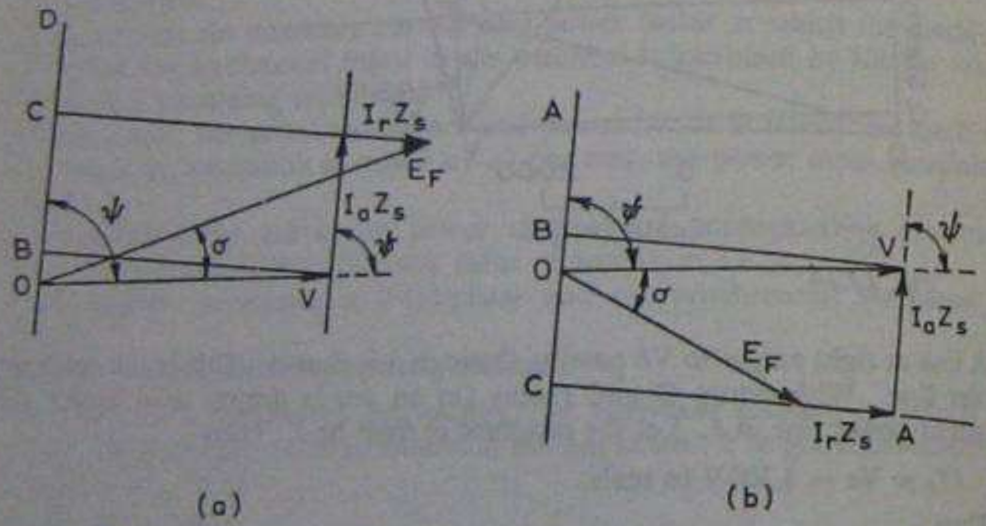


Fig. 12.15 POWER TRANSFER FOR A SYNCHRONOUS MACHINE
(*a*) Generator (*b*) Motor

the corresponding complexor diagram for motor operation also at a lagging power factor. The power transfer is

$$P = 3VI\cos\phi = 3VI_a \tag{12.33}$$

where V is the phase voltage and I is the phase current.

The projection of the complexors of Fig. 12.15(*a*) on the steady-state limit of stability line OD gives

$$I_a Z_s = E_F \cos(\psi - \sigma) - V \cos\psi \tag{12.34}$$

Substituting the expression for I_a obtained from eqn. (12.34) in eqn. (12.33) gives

$$P = \frac{3V}{Z_s}\{E_F \cos(\psi - \sigma) - V \cos\psi\} \tag{12.35}$$

Following the same procedure for motor action and using Fig. 12.15 (*b*) the power transfer is found to be

$$P = \frac{3V}{Z_s}\{V \cos\psi - E_F \cos(\psi + \sigma)\} \tag{12.36}$$

Evidently eqn. (12.36) will cover both generator action and motor action if the power transfer P and the load angle σ are taken, conventionally, to be positive for generator action and negative for motor action.

Since, for steady-state operation, the speed of a synchronous machine is constant, the torque developed is

$$T = \frac{P}{2\pi n_0} = \frac{3}{2\pi n_0}\frac{V}{Z_s}\{E_F \cos(\psi - \sigma) - V \cos\psi\} \tag{12.37}$$

In many synchronous machines $X_s \gg R$, in which case $Z_s\underline{/\psi} \approx X_s\underline{/90^\circ}$. When this approximation is permissible eqn. (12.35) becomes

$$P = \frac{3V}{Z_s}\{E_F \cos(90^\circ - \sigma) - V \cos 90^\circ\}$$

$$= \frac{3VE_F}{X_s}\sin\sigma \tag{12.38}$$

Similarly eqn. (12.37) becomes

$$T = \frac{3}{2\pi n_0}\frac{VE_F}{X_s}\sin\sigma \tag{12.39}$$

The power/load-angle (or torque/load-angle) characteristic is shown in Fig. 12.16. The dotted parts of this characteristic refer to operation beyond the steady-state limit of stability. Usually stable operation cannot be obtained beyond this limit, so that if the load angle exceeds $\pm 90°$ the operation is dynamic with the machine either

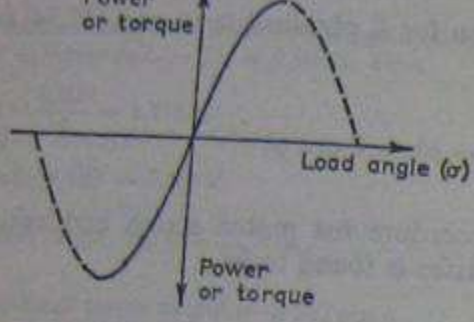


Fig. 12.16 POWER/LOAD-ANGLE AND TORQUE/LOAD-ANGLE CHARACTERISTICS OF A SYNCHRONOUS MACHINE CONNECTED TO INFINITE BUSBARS

accelerating or decelerating. In this case eqns. (12.38) and (12.39) are only approximately true.

12.16 Synchronizing Power and Synchronizing Torque Coefficients

A synchronous machine, whether a generator or a motor, when synchronized to infinite busbars has an inherent tendency to remain synchronized.

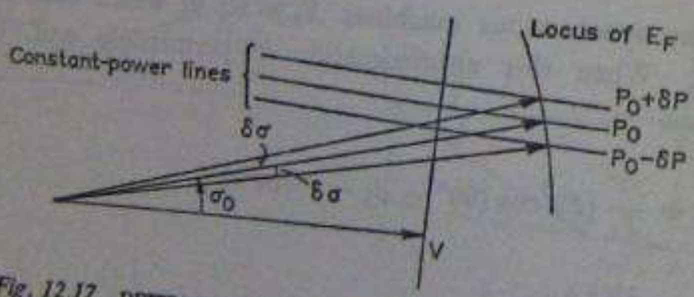


Fig. 12.17 DETERMINATION OF SYNCHRONIZING POWER COEFFICIENT

In Fig. 12.17, which applies to generator operation at a lagging power factor, the complexor diagram is part of the general load diagram. At a steady load angle σ_0 the steady power transfer is P_0.

Suppose that, due to a transient disturbance, the rotor of the machine accelerates, so that the load angle increases by $\delta\sigma$. This alters the operating point of the machine to a new constant-power line and the load on the machine increases to $P_0 + \delta P$. Since the steady power input remains unchanged, this additional load retards the machine and brings it back to synchronism.

Similarly, if owing to a transient disturbance, the rotor decelerates so that the load angle decreases, the load on the machine is thereby reduced to $P_0 - \delta P$. This reduction in load causes the rotor to accelerate and the machine is again brought back to synchronism.

Clearly the effectiveness of this inherent correcting action depends on the extent of the change in power transfer for a given change in load angle. A measure of this effectiveness is given by the *synchronizing power coefficient*, which is defined as

$$P_s = \frac{dP}{d\sigma} \tag{12.40}$$

From eqn. (12.35),

$$P = \frac{3V}{Z_s}\{E_F \cos(\psi - \sigma) - V \cos\psi\} \tag{12.35}$$

so that

$$P_s = \frac{dP}{d\sigma} = \frac{3VE_F}{Z_s}\sin(\psi - \sigma) \tag{12.41}$$

Similarly the synchronizing torque coefficient is defined as

$$T_s = \frac{dT}{d\sigma} = \frac{1}{2\pi n_0}\frac{dP}{d\sigma} \tag{12.42}$$

From eqn. (12.42), therefore,

$$T_s = \frac{3}{2\pi n_0}\frac{VE_F}{Z_s}\sin(\psi - \sigma) \tag{12.43}$$

In many synchronous machines $X_s \gg R$, in which case eqns. (12.42) and (12.43) become

$$P_s = \frac{3VE_F}{X_s}\cos\sigma \tag{12.44}$$

$$T_s = \frac{3}{2\pi n_0}\frac{VE_F}{X_s}\cos\sigma \tag{12.45}$$

Eqns. (12.44) and (12.45) show that the restoring action is greatest when $\sigma = 0$, i.e. on no-load. The restoring action is zero when $\sigma = \pm 90°$. At these values of load angle the machine would be at the steady-state limit of stability and in a condition of unstable equilibrium. It is impossible, therefore, to run a machine at the steady-state limit of stability since its ability to resist small changes is zero unless the machine is provided with a special fast-acting excitation system.

EXAMPLE 12.5 A 2 MVA 3-phase 8-pole alternator is connected to 6,000 V 50 Hz busbars and has a synchronous reactance of 4 Ω per phase. Calculate the synchronizing power and synchronizing torque per mechanical degree of rotor displacement at no-load. (Assume normal excitation.)

The synchronizing power coefficient is

$$P_s = \frac{3VE_f}{X_s}\cos\sigma \tag{12.44}$$

On no-load the load angle $\sigma = 0$.

Since there are 4 pole-pairs, 1 mechanical degree of displacement is equivalent to 4 electrical degrees; therefore

$$P_s = 3 \times \frac{6{,}000}{\sqrt{3}} \times \frac{6{,}000}{\sqrt{3} \times 4} \times \frac{4}{1{,}000} \times \frac{\pi}{180} = \underline{627\,\text{kW/mech. deg}}$$

Synchronous speed of alternator, $n_0 = \dfrac{f}{p} = 12{\cdot}5$ rev/s

Thus

$$2\pi n_0 T_s = 627 \times 10^3$$

and

Synchronizing torque, $T_s = \underline{8{,}000\,\text{N-m/mech. deg}}$

12.17 Oscillation of Synchronous Machines

In the previous sections, transient accelerations or decelerations of an alternator rotor were assumed in order to investigate the synchronizing power and synchronizing torque. Such transients may be caused by irregularities in the driving torque of the prime mover or, in the case of a motor, by irregularities in the load torque, or by irregularities in other machines connected in parallel, or by sudden changes in load.

Normally the inherent stability of alternators when running in parallel quickly restores the steady-state condition, but if the effect is sufficiently marked, the machine may break from synchronism. Moreover, if the disturbance is cyclic in effect, recurring at regular intervals, it will produce forced oscillations in the machine rotor. If the frequency of this cyclic disturbance approaches the value of the natural frequency of the rotor, when connected to the busbar system, the rotor may be subject to continuous oscillation and may eventually break from synchronism. This continuous oscillation of the rotor (periods of acceleration and deceleration) is sometimes known as *phase swinging* or *hunting*.

Fig. 12.18 shows the torque/load-angle characteristic of a synchronous generator. The steady input torque is T_0, corresponding to a steady-state load angle σ_0. Suppose a transient disturbance occurs such as to make the rotor depart from the steady state by σ'. Let σ' be sufficiently small to assume that the synchronizing

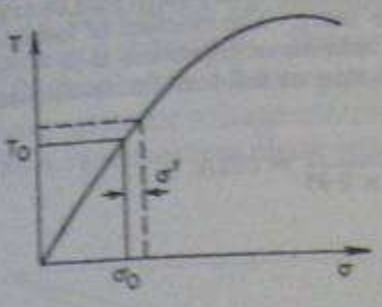


Fig. 12.18 OSCILLATION OF A SYNCHRONOUS MACHINE CONNECTED TO INFINITE BUSBARS

torque is constant; i.e. the torque/load-angle characteristic is assumed to be linear over the range of σ' considered.

Let T_s = Synchronizing torque coefficient (*N-m*/mech. rad)
σ' = Load angle deviation from steady-state position (mech. rad)
J = Moment of inertia of rotating system ($kg\text{-}m^2$)

Assuming that there is no damping,

$$J\frac{d^2\sigma'}{dt^2} = -T_s\sigma' \tag{12.46}$$

The solution of this differential equation is

$$\sigma' = \sigma_m' \sin\left(\sqrt{\frac{T_s}{J}}\,t + \psi\right) \tag{12.47}$$

From eqn. (12.47), the frequency of undamped oscillation is

$$f = \frac{1}{2\pi}\sqrt{\frac{T_s}{J}} \tag{12.48}$$

Synchronous machines intended for operation on infinite busbars are provided with damping windings in order to prevent the sustained oscillations predicted by eqn. (12.48).

In salient-pole machines the damping winding takes the form of a short-circuited cage consisting of copper bars of relatively large cross-section embedded in the rotor pole-face. In cylindrical-rotor machines the solid rotor provides considerable damping, but a cage winding may also be provided. This consists of copper fingers inserted in the rotor slots below the slot wedges and joined together at each end of the rotor. The currents induced in the damping bars give a damping torque which prevents continuous oscillation of the rotor.

EXAMPLE 12.6 A 3-phase 3·3 kV 2-pole 3,000 rev/min 934 kW synchronous motor has an efficiency of 0·95 p.u. and delivers full-load torque with its excitation adjusted so that the input power factor is unity. The moment of inertia of the motor and its load is 30 kg-m², and its synchronous impedance is $(0 + j11\cdot1)\Omega$. Determine the period of undamped oscillation on full-load for small changes in load angle.

$$\text{Input current, } I = \frac{934 \times 10^3}{\sqrt{3} \times 3\cdot3 \times 10^3 \times 0\cdot95} = 172\,\text{A}$$

Taking the phase voltage as reference,

$$E_F = V - IX_s$$

$$= \frac{3\cdot3 \times 10^3}{\sqrt{3}}\underline{/0^\circ} - (172\underline{/0^\circ} \times 11\cdot1\underline{/90^\circ}) = 2{,}700\underline{/-45^\circ}\,\text{V}$$

The synchronizing torque coefficient is

$$T_s = \frac{3}{2\pi n_0}\frac{VE_F}{X_s}\cos\sigma \qquad (12.45)$$

$$= \frac{3}{2\pi 50} \times \frac{3\cdot3 \times 10^3}{\sqrt{3}} \times \frac{2{,}700}{11\cdot1} \times 0\cdot707 = 3\cdot14 \times 10^3\,\text{N-m/rad}$$

The undamped frequency of oscillation is

$$f = \frac{1}{2\pi}\sqrt{\frac{T_s}{J}}$$

The period of oscillation is

$$T = \frac{1}{f} = 2\pi\sqrt{\frac{30}{3\cdot14 \times 10^3}} = \underline{\underline{0\cdot612\,\text{s}}}$$

12.18 Synchronous Motors

A synchronous motor will not develop a driving torque unless it is running at synchronous speed, since at any other speed the field poles will alternately be acting on the effective N and S poles of the rotating field and only a pulsating torque will be produced. For starting either (*a*) the induction motor principle or (*b*) a separate starting motor must be used. If the latter method is used the machine must be run up to synchronous speed and synchronized as an alternator. To obviate this trouble, synchronous motors are usually started as induction motors, and have a squirrel-cage winding embedded in the rotor pole faces to give the required action. When the machine has run up to almost synchronous speed the d.c. excitation is switched on to the rotor, and it then pulls into synchronism. The induction motor action then ceases (see Chapter 13).

The starting difficulties of a synchronous motor severely limit its usefulness—it may only be used where the load may be reduced for starting and where starting is infrequent. Once started, the motor has the advantage of running at a constant speed with any desired power factor. Typical applications of synchronous motors are the driving of ventilation or pumping machinery where the machines run almost continuously. Synchronous motors are often run with no load to utilize their leading power factor characteristic for power factor correction or voltage control. In these applications the machine is called a synchronous phase modifier.

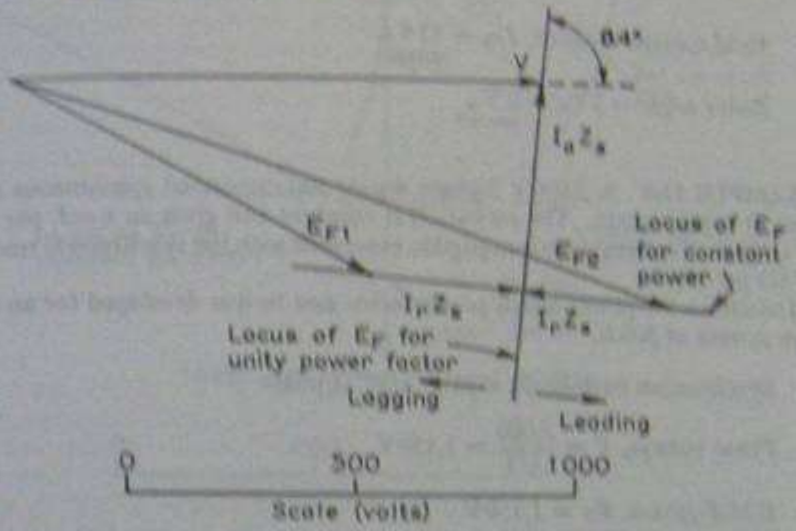


Fig. 12.19

EXAMPLE 12.7 A 2,000 V 3-phase 4-pole star-connected synchronous machine has resistance and synchronous reactance per phase of 0·2 Ω and 1·9 Ω respectively.

Calculate the e.m.f. and the rotor displacement when the machine acts as a motor with an input of 800 kW at power factors of 0·8 lagging and leading.

If a field current of 40 A is required to produce an e.m.f. per phase equal to rated phase voltage, determine also the field current for each condition.

Synchronous impedance, $Z_s = 0{\cdot}2 + j1{\cdot}9 = 1{\cdot}91\underline{/84^\circ}\,\Omega$/phase

Constant phase terminal voltage, $V = \dfrac{2{,}000}{\sqrt{3}} = 1{,}150\,\text{V}$

Total phase current in both cases $= \dfrac{800 \times 10^3}{\sqrt{3} \times 2{,}000 \times 0{\cdot}8} = 288\,\text{A}$

Active component of current in both cases, $I_a = 288 \times 0{\cdot}8 = 230\,\text{A}$

Reactive component of current in both cases, $I_r = 288 \times 0{\cdot}6 = 173\,\text{A}$

$I_aZ_s = 230 \times 1{\cdot}91 = 440\,\text{V}$

$I_rZ_s = 173 \times 1{\cdot}91 = 330\,\text{V}$

Fig. 12.19 is now drawn to scale for the motoring condition.

At the lagging power factor the excitation voltage is measured from the complexor diagram as E_{F1} = 880 V/phase.

$$\text{Field current required, } I_{F1} = 40 \times \frac{880}{1{,}150} = 30{\cdot}5\,\text{A}$$

The rotor displacement is the phase angle between E_{F1} and V with the rotor lagging for motor action as previously described. Therefore at the lagging p.f.

Rotor angle = $27°_e$ = $13{\cdot}5°_m$ for a 4-pole machine

At the leading p.f. the excitation voltage, E_{F2} = 1,520 V/phase

Field current required, I_{F2} = 52·9 A

Rotor angle = $17°_e$ = $8{\cdot}5°_m$

EXAMPLE 12.8 A 2,000 V 3-phase 4-pole star-connected synchronous motor runs at 1,500 rev/min. The excitation is constant and gives an e.m.f. per phase of 1,150 V. The resistance is negligible compared with the synchronous reactance of 3 Ω per phase.

Determine the power input, power factor and torque developed for an armature current of 200 A.

Synchronous impedance = $j3 = 3\angle 90°$ Ω/phase

$$\text{Phase voltage, } V = \frac{2{,}000}{\sqrt{3}} = 1{,}150\,\text{V}$$

E.M.F./phase, E_F = 1,150 V

$IZ_s = 200 \times 3 = 600$ V

In Fig. 12.20 V represents the phase voltage taken as reference complexor.

A circular arc whose radius represents the open-circuit voltage of 1,150 V is the locus of E_F for constant excitation.

AB is the locus of E_F for unity power factor operation; in this case AB is perpendicular to V since the phase angle of Z is 90°.

A circle whose radius represents 600 V is the locus of E_F for constant kVA operation. For the actual operating conditions E_F must lie at the intersection of the two circles.

From the diagram,

$I_a Z_s$ = 580 V

Active component of current, I_a = 193 A

Therefore

$$\text{Total power input} = \frac{3VI_a}{1{,}000} = \frac{3 \times 1{,}150 \times 193}{1{,}000} = 666\,\text{kW}$$

$$\text{Operating power factor} = \frac{I_a}{I} = \frac{193}{200} = 0{\cdot}96 \text{ lagging}$$

$$\text{Torque developed, } T = \frac{3VI_a}{2\pi n_s}$$

$$= \frac{3 \times 1150 \times 193}{2\pi \times 1{,}500} \times 60$$

$$= \underline{4{,}250 \text{ N-m}}$$

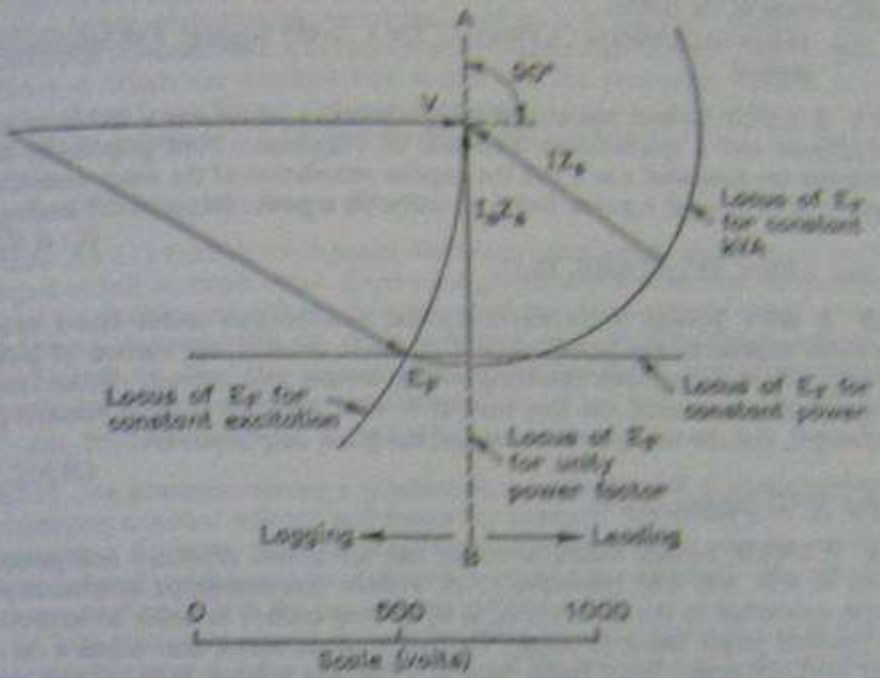


Fig. 12.20

PROBLEMS

12.1 A 3-phase 11kV star-connected alternator has an effective armature resistance of 1Ω and a synchronous reactance of 20Ω per phase. Calculate the percentage regulation for a load of 1,500kW at p.f.s of (*a*) 0·8 lagging, (*b*) unity, (*c*) 0·8 leading.

Ans. 22 per cent, 4·25 per cent, −13·4 per cent.

12.2 Describe the tests carried out in order that the synchronous impedance of an alternator can be obtained. By means of diagrams show how the synchronous impedance can be used to determine the regulation of an alternator at a particular load and power factor.

A 6,600V 3-phase star-connected alternator has a synchronous impedance of $(0{\cdot}4 + j6)$Ω/phase. Determine the percentage regulation of the machine when supplying a load of 1,000kW at normal voltage and p.f. (i) 0·866 lagging, (ii) unity, (iii) 0·866 leading, giving complexor diagrams in each case. (*H.N.C.*)

Ans. 9·7 per cent, 1·84 per cent, −6·03 per cent.

Learning Outcome 4.4

Predict the values of power, stator current, power factor and load angle for specified values of torque and excitation given the synchronous impedance and excitation values

17.2 Starting a synchronous motor

A synchronous motor cannot start by itself; consequently, the rotor is usually equipped with a squirrel-cage winding so that it can start up as an induction motor. When the stator is connected to the 3-phase line, the motor accelerates until it reaches a speed slightly below synchronous speed. The dc excitation is suppressed during this starting period.

While the rotor accelerates, the rotating flux created by the stator sweeps across the slower moving salient poles. Because the coils on the rotor possess a relatively large number of turns, a high voltage is induced in the rotor winding when it turns at low speeds. This voltage appears between the slip-rings and it decreases as the rotor accelerates, eventually becoming negligible when the rotor approaches synchronous speed. To limit the voltage, and to improve the starting torque, we either short-circuit the slip-rings or connect them to an auxiliary resistor during the starting period.

If the power capacity of the supply line is limited, we sometimes have to apply reduced voltage to the stator. As in the case of induction motors, we use either autotransformers or series reactors to limit the starting current (see Chapter 20). Very large synchronous motors (20 MW and more) are sometimes brought up to speed by an auxiliary motor, called a *pony motor*. Finally, in some big installations the motor may be brought up to speed by a variable-frequency electronic source.

17.3 Pull-in torque

As soon as the motor is running at close to synchronous speed, the rotor is excited with dc current. This produces alternate N and S poles around the circumference of the rotor (Fig. 17.5). If the poles at this moment happen to be facing poles of *opposite* polarity on the stator, a strong magnetic attraction is set up between them. The mutual attraction locks the rotor and stator poles together, and the rotor is literally yanked into step with the revolving field. The torque developed at this moment is appropriately called the *pull-in torque*.

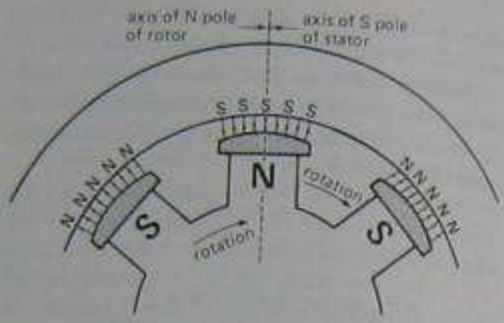


Figure 17.5
The poles of the rotor are attracted to the opposite poles on the stator. At no-load the axes of the poles coincide.

The pull-in torque of a synchronous motor is powerful, but the dc current must be applied at the right moment. For example, if it should happen that the emerging N, S poles of the rotor are opposite the N, S poles of the stator, the resulting magnetic repulsion produces a violent mechanical shock. The motor will immediately slow down and the circuit breakers will trip. In practice, starters for synchronous motors are designed to detect the precise moment when excitation should be applied. The motor then pulls automatically and smoothly into step with the revolving field.

Once the motor turns at synchronous speed, no voltage is induced in the squirrel-cage winding and so it carries no current. Consequently, the behavior of a synchronous motor is entirely different from that of an induction motor. Basically, a synchronous motor rotates because of the magnetic attraction between the poles of the rotor and the opposite poles of the stator.

To reverse the direction of rotation, we simply interchange any two lines connected to the stator.

17.4 Motor under load—general description

When a synchronous motor runs at no-load, the rotor poles are directly opposite the stator poles and their axes coincide (Fig. 17.5). However, if we apply a mechanical load, the rotor poles fall slightly

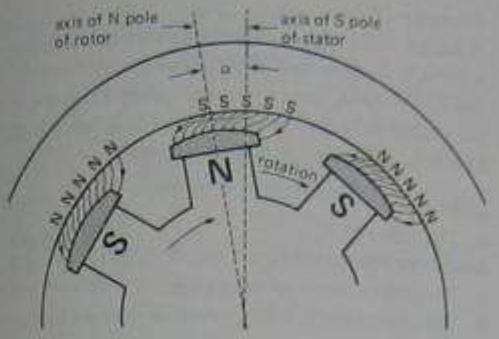


Figure 17.6
The rotor poles are displaced with respect to the axes of the stator poles when the motor delivers mechanical power.

behind the stator poles, but the rotor continues to turn at synchronous speed. The mechanical angle α between the poles increases progressively as we increase the load (Fig. 17.6). Nevertheless, the magnetic attraction keeps the rotor locked to the revolving field, and the motor develops an ever more powerful torque as the angle increases.

But there is a limit. If the mechanical load exceeds the *pull-out torque* of the motor, the rotor poles suddenly pull away from the stator poles and the motor comes to a halt. A motor that pulls out of step creates a major disturbance on the line, and the circuit breakers immediately trip. This protects the motor because both the squirrel-cage and stator windings overheat rapidly when the machine ceases to run at synchronous speed.

The pull-out torque depends upon the magnetomotive force developed by the rotor and the stator poles. The mmf of the rotor poles depends upon the dc excitation I_x, while that of the stator depends upon the ac current flowing in the windings. The pull-out torque is usually 1.5 to 2.5 times the nominal full-load torque.

The mechanical angle α between the rotor and stator poles has a direct bearing on the stator current. As the angle increases, the current increases. This is to be expected because a larger angle corresponds to a bigger mechanical load, and the increased power can only come from the 3-phase ac source.

17.5 Motor under load—simple calculations

We can get a better understanding of the operation of a synchronous motor by referring to the equivalent circuit shown in Fig. 17.7a. It represents one phase of a wye-connected motor. It is identical to the equivalent circuit of an ac generator, because both machines are built the same way. Thus, the flux Φ created by the rotor induces a voltage E_o in the stator. This flux depends on the dc exciting current I_x. Consequently, E_o varies with the excitation.

As already mentioned, the rotor and stator poles are lined up at no-load. Under these conditions, induced voltage E_o is in phase with the line-to-neutral voltage E (Fig. 17.7b). If, in addition, we adjust the excitation so that $E_o = E$, the motor "floats" on the line and the line current I is practically zero. In effect, the only current needed is to supply the small windage and friction losses in the motor, and so it is negligible.

What happens if we apply a mechanical load to the shaft? The motor will begin to slow down, causing the rotor poles to fall behind the stator poles by an angle α. Due to this mechanical shift, E_o reaches its maximum value a little later than before. Thus, referring to Fig. 17.7c, E_o is now δ electrical degrees behind E. The mechanical displacement α produces an *electrical* phase shift δ between E_o and E.

The phase shift produces a difference of potential E_x across the synchronous reactance X_s given by

$$E_x = E - E_o$$

Consequently, a current I must flow in the circuit, given by

$$jIX_s = E_x$$

from which

$$I = -jE_x/X_s$$
$$= -j(E - E_o)/X_s$$

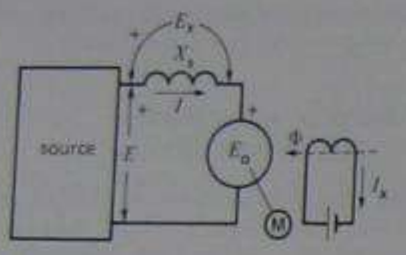


Figure 17.7a
Equivalent circuit of a synchronous motor, showing one phase.

Figure 17.7b
Motor at no-load, with E_o adjusted to equal E.

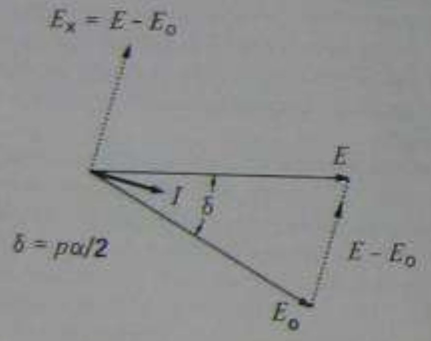


Figure 17.7c
Motor under load E_o has the same value as in Fig. 17.7b, but it lags behind E.

The current lags 90° behind E_x because X_s is inductive. The phasor diagram under load is shown in Fig. 17.7c. Because I is nearly in phase with E, the motor absorbs active power. This power is entirely transformed into mechanical power, except for the relatively small copper and iron losses in the stator.

In practice, the excitation voltage E_o is adjusted to be greater or less than the supply voltage E. Its value depends upon the power output of the motor and the desired power factor.

Example 17-2a

A 500 hp, 720 r/min synchronous motor connected to a 3980 V, 3-phase line generates an excitation voltage E_o of 1790 V (line-to-neutral) when the dc exciting current is 25 A. The synchronous reactance is 22 Ω and the torque angle between E_o and E is 30°.

Calculate

a. The value of E_x
b. The ac line current
c. The power factor of the motor
d. The approximate horsepower developed by the motor
e. The approximate torque developed at the shaft

Solution

This problem can best be solved by using vector notation.

a. The voltage E (line-to-neutral) applied to the motor has a value

$$E = E_L/\sqrt{3} = 3980/\sqrt{3}$$
$$= 2300 \text{ V}$$

Let us select E as the reference phasor, whose angle with respect to the horizontal axis is assumed to be zero. Thus,

$$E = 2300\angle 0°$$

It follows that E_o is given by the phasor

$$E_o = 1790\angle -30°$$

The equivalent circuit per phase is given in Fig. 17.8a.

Moving clockwise around the circuit and applying Kirchhoff's voltage law we can write

$$-E + E_x + E_o = 0$$
$$E_x = E - E_o$$
$$= 2300\angle 0° - 1790\angle -30°$$
$$= 2300 (\cos 0° + j \sin 0°) - 1790 (\cos -30° + j \sin -30°)$$
$$= 2300 - 1550 + j\,895$$
$$= 750 + j\,895$$
$$= 1168\angle 50°$$

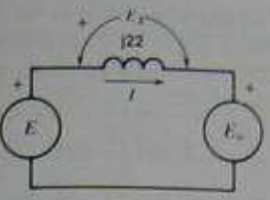


Figure 17.8a
Equivalent circuit of a synchronous motor connected to a source E.

Thus, phasor E_x has a value of 1168 V and it leads phasor E by 50°.

b. The line current I is given by

$$j22\,I = E_x$$
$$I = \frac{1168 \angle 50°}{22 \angle 90°}$$
$$= 53\angle -40°$$

Thus, phasor I has a value of 53 A and it lags 40° behind phasor E.

c. The power factor of the motor is given by the cosine of the angle between the line-to-neutral voltage E across the motor terminals and the current I. Hence,

$$\text{power factor} = \cos\theta = \cos 40°$$
$$= 0.766, \text{ or } 76.6\%$$

The power factor is lagging because the current lags behind the voltage.
The complete phasor diagram is shown in Fig. 17.8b.

d. Total active power input to the stator:

$$P_s = 3 \times E_{LN} I_L \cos\theta$$
$$= 3 \times 2300 \times 53 \times \cos 40°$$
$$= 280\,142 \text{ W} = 280.1\text{kW}$$

Neglecting the I^2R losses and iron losses in the stator, the electrical power transmitted across the airgap to the rotor is 280.1 kW.
Approximate horsepower developed:

$$P = 280.1 \times 10^3/746 = 375 \text{ hp}$$

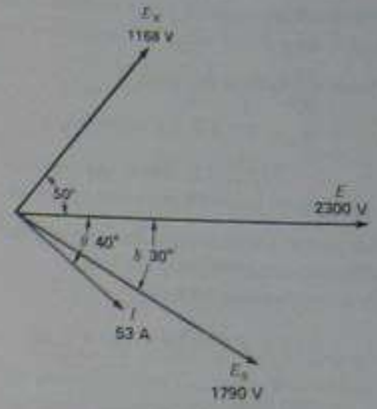


Figure 17.8b
See Example 17-2.

e. Approximate torque:

$$T = \frac{9.55 \times P}{n} = \frac{9.55 \times 280.1 \times 10^3}{720}$$
$$= 3715 \text{ N·m}$$

Example 17-2b

The motor in Example 17-2a has a stator resistance of 0.64 Ω per phase and possesses the following losses:

I^2R losses in the rotor:	3.2 kW
Stator core loss:	3.3 kW
Windage and friction loss:	1.5 kW

Calculate

a. The actual horsepower developed
b. The actual torque developed at the shaft
c. The efficiency of the motor

Solution

a. Power input to the stator is 280.1 kW

Stator I^2R losses $= 3 \times 53^2 \times 0.64\ \Omega = 5.4$ kW
Total stator losses $= 5.4 + 3.3 = 8.7$ kW
Power transmitted to the rotor $= 280.1 - 8.7$ $= 271.4$ kW

The power at the shaft is the power to the rotor minus the windage and friction losses. The rotor

I^2R losses are supplied by an external dc source and so they do not affect the mechanical power.

Power available at the shaft:

$$P_o = 271.4 - 1.5 = 269.9 \text{ kW}$$
$$= \frac{269.9 \times 10^3}{746} = 361.8 \text{ hp}$$

This power is very close to the approximate value calculated in Example 17-2a.

b. The corresponding torque is:

$$T = \frac{9.55 \times P}{n} = \frac{9.55 \times 269.9 \times 10^3}{720}$$
$$= 3580 \text{ N·m}$$

c. Total losses $= 5.4 + 3.3 + 3.2 + 1.5 = 13.4$ kW

Total power input $= 280.1 + 3.2 = 283.3$ kW
Total power output $= 269.9$ kW
Efficiency $= 269.9/283.3 = 0.9527 = 95.3$ %

Note that the stator resistance of 0.64 Ω is very small compared to the reactance of 22 Ω. Consequently, the true phasor diagram is very close to the phasor diagram of Fig. 17.8b.

17.6 Power and torque

When a synchronous motor operates under load, it draws active power from the line. The power is given by the same equation we previously used for the synchronous generator in Chapter 16:

$$P = (E_oE/X_s) \sin \delta \qquad (16.5)$$

As in the case of a generator, the active power absorbed by the motor depends upon the supply voltage E, the excitation voltage E_o, and the phase angle δ between them. If we neglect the relatively small I^2R and iron losses in the stator, all the power is transmitted across the air gap to the rotor. This is analogous to the power P_r transmitted across the air gap of an induction motor (Section 13.13). However, in a synchronous motor, the rotor I^2R losses are entirely supplied by the dc source. Consequently, all the power transmitted across the air gap is available in the form of mechanical power. The mechanical power developed by a synchronous motor is therefore expressed by the equation

$$P = \frac{E_oE}{X_s} \sin \delta \qquad (17.2)$$

where

P = mechanical power of the motor, per phase [W]
E_o = line-to-neutral voltage induced by I_x [V]
E = line-to-neutral voltage of the source [V]
X_s = synchronous reactance per phase [Ω]
δ = torque angle between E_o and E [electrical degrees]

This equation shows that the mechanical power increases with the torque angle, and its maximum value is reached when δ is 90°. The poles of the rotor are then midway between the N and S poles of the stator. The peak power P_{max} (per-phase) is given by

$$P_{max} = \frac{E_oE}{X_s} \qquad (17.3)$$

As far as torque is concerned, it is directly proportional to the mechanical power because the rotor speed is fixed. The torque is derived from Eq. 3.5:

$$T = \frac{9.55\, P}{n_s} \qquad (17.4)$$

where

T = torque, per phase [N·m]
P = mechanical power, per phase [W]
n_s = synchronous speed [r/min]
9.55 = a constant [exact value = $60/2\pi$]

The maximum torque the motor can develop is called the pull-out torque, mentioned previously. It occurs when $\delta = 90°$ (Fig. 17.9).*

* The remarks in this section apply to motors having smooth rotors. Most synchronous motors have salient poles; in this case the pull-out torque occurs at an angle of about 70°.

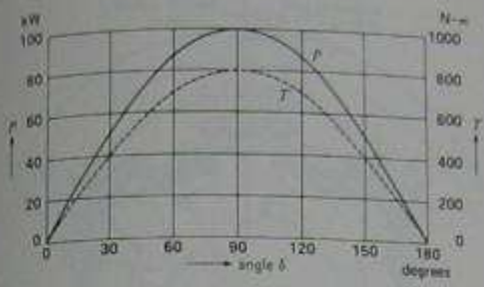


Figure 17.9
Power and torque per phase as a function of the torque angle δ. Synchronous motor rated 150 kW (200 hp), 1200 r/min, 3-phase, 60 Hz. See Example 17-3.

Example 17-3

A 150 kW, 1200 r/min, 460 V, 3-phase synchronous motor has a synchronous reactance of 0.8 Ω, per phase. If the excitation voltage E_o is fixed at 300 V, per phase, determine the following:

a. The power versus δ curve
b. The torque versus δ curve
c. The pull out torque of the motor

Solution

a. The line-to-neutral voltage is

$$E = E_L/\sqrt{3} = 460/\sqrt{3} = 266 \text{ V}$$

The mechanical power per phase is

$$P = (E_o E/X_s)\sin\delta \qquad (17.2)$$
$$= (266 \times 300/0.8)\sin\delta$$
$$= 99\,750 \sin\delta \text{ [W]}$$
$$= 100 \sin\delta \text{ [kW]}$$

By selecting different values for δ, we can calculate the corresponding values of P and T, per phase.

δ [°]	*P* [kW]	*T* [N·m]
0	0	0
30	50	400
60	86.6	693
90	100	800
120	86.6	693
150	50	400
180	0	0

(continued)

These values are plotted in Fig. 17.9.

b. The torque curve can be found by applying Eq. 17.4:

$$T = 9.55\, P/n_s$$
$$= 9.55\, P/1200$$
$$= P/125$$

c. The pull-out torque T_{max} coincides with the maximum power output:

$$T_{max} = 800 \text{ N·m}$$

The actual pull-out torque is 3 times as great (2400 N·m) because this is a 3-phase machine. Similarly, the power and torque values given in Fig. 17.9 must also be multiplied by 3. Consequently, this 150 kW motor can develop a maximum output of 300 kW, or about 400 hp.

17.7 Mechanical and electrical angles

As in the case of synchronous generators, there is a precise relationship between the mechanical angle α, the torque angle δ and the number of poles p. It is given by

$$\delta = p\alpha/2 \qquad (17.5)$$

Example 17-4

A 3-phase, 6000 kW, 4 kV, 180 r/min, 60 Hz motor has a synchronous reactance of 1.2 Ω. At full-load the rotor poles are displaced by a *mechanical* angle of 1° from their no-load position. If the line-to-neutral excitation E_o = 2.4 kV, calculate the mechanical power developed.

Solution

The number of poles is

$$p = 120\, f/n_s = 120 \times 60/180 = 40$$

The electrical torque angle is

$$\delta = p\alpha/2 = (40 \times 1)/2 = 20°$$

Assuming a wye connection, the voltage E applied to the motor is

$$E = E_L/\sqrt{3} = 4\text{ kV}/\sqrt{3}$$
$$= 2.3\text{ kV}$$
$$= 2309\text{ V}$$

and the excitation voltage is

$$E_o = 2400\text{ V}$$

The mechanical power developed per phase is

$$P = (E_oE/X_s)\sin\delta \qquad (17.2)$$
$$= (2400 \times 2309/1.2)\sin 20°$$
$$= 1\,573\,300$$
$$= 1573\text{ kW}$$

$$\text{Total power} = 3 \times 1573$$
$$= 4719\text{ kW } (\sim 6300\text{ hp})$$

17.8 Reluctance torque

If we gradually reduce the excitation of a synchronous motor when it is running at no-load, we find that the motor continues to run at synchronous speed even when the exciting current is zero. The reason is that the flux produced by the stator prefers to cross the short gap between the salient poles and the stator rather than the much longer air gap between the poles. In other words, because the reluctance of the magnetic circuit is less in the axis of the salient poles, the flux is concentrated as shown in Fig. 17.10a. On account of this phenomenon, the motor develops a *reluctance torque*.

If a mechanical load is applied to the shaft, the rotor poles will fall behind the stator poles, and the stator flux will have the shape shown in Fig. 17.10b. Thus, a considerable reluctance torque can be developed without any dc excitation at all.

The reluctance torque becomes zero when the rotor poles are midway between the stator poles. The reason is that the N and S poles on the stator attract the salient poles in opposite directions (Fig. 17.10c). Consequently, the reluctance torque is zero precisely at that angle where the regular torque T attains its maximum value, namely at $\delta = 90°$.

Fig. 17.11 shows the reluctance torque as a function of the angle δ. The torque reaches a maximum positive value at $\delta = 45°$. For larger angles it attains a maximum *negative* value at $\delta = 135°$. Obviously, to run as a reluctance-torque motor, the angle must lie between zero and 45°. Although a positive torque is still developed between 45° and 90°, this is an unstable region of operation. The reason is that as the angle increases the power decreases.

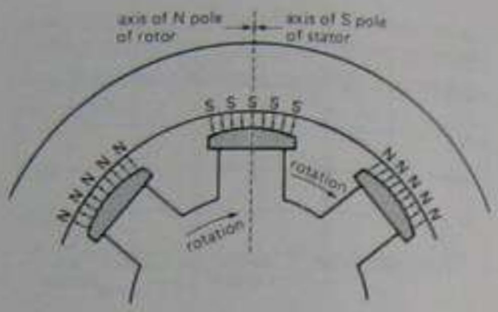


Figure 17.10a
The flux produced by the stator flows across the air gap through the salient poles.

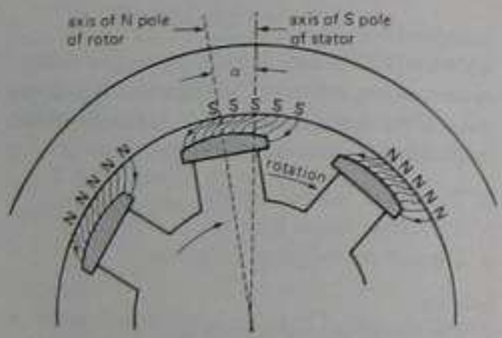


Figure 17.10b
The salient poles are attracted to the stator poles, thus producing a reluctance torque.

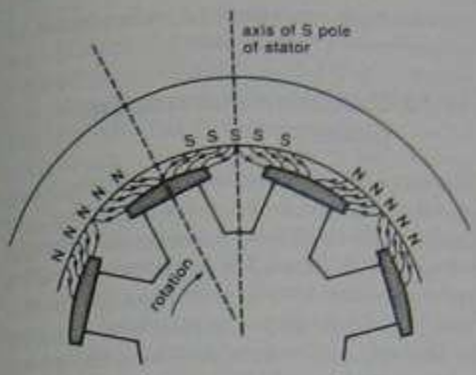


Figure 17.10c
The reluctance torque is zero when the salient poles are midway between the stator poles.

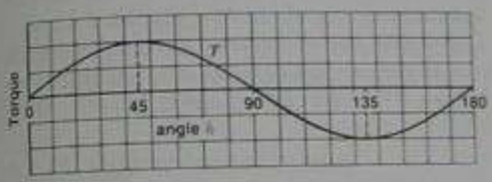


Figure 17.11
Reluctance torque versus the torque angle.

As in the case of a conventional synchronous motor, the mechanical power curve has exactly the same shape as the torque curve. Thus, in the absence of dc excitation, the mechanical power reaches a peak at $\delta = 45°$.

Does the saliency of the poles modify the power and torque curves shown in Fig. 17.9? The answer is yes. In effect, the curves shown in Fig. 17.9 are those of a smooth-rotor synchronous motor. The torque of a salient-pole motor is equal to the sum of the smooth-rotor component and the reluctance-torque component of Fig. 17.11. Thus, the true torque curve of a synchronous motor has the shape (3) given in Fig. 17.12.

The peak reluctance torque is about 25 percent of the peak *smooth-rotor torque*. As a result, the peak torque of a salient-pole motor is about 8 percent greater than that of a smooth-rotor motor, as

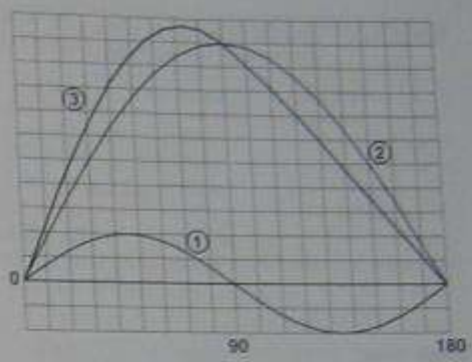


Figure 17.12
In a synchronous motor the reluctance torque (1) plus the smooth-rotor torque (2) produce the resultant torque (3). Torque (2) is due to the dc excitation of the rotor.

can be seen in Fig. 17.12. However, the difference is not very great, and for this reason we shall continue to use Eqs. 17.2 and 17.5 to describe synchronous motor behavior.

17.9 Losses and efficiency of a synchronous motor

In order to give the reader a sense of the order of magnitude of the pull-out torque, resistance, reactance, and losses of a synchronous motor, we have drawn up Table 17A. It shows the characteristics of a 2000 hp and a 200 hp synchronous motor, respectively labeled Motor A and Motor B.

The following points should be noted:

1. The torque angle at full-load ranges between 27° and 37°. It corresponds to the electrical angle δ mentioned previously.
2. The power needed to excite the 2000 hp motor (4.2 kW) is only about twice that needed for the 200 hp motor (2.1 kW). In general, the larger the synchronous motor the smaller is the per-unit power needed to excite it.
3. The total losses of Motor A (38 kW) are only four times those of Motor B (9.5 kW) despite the

(a)

(b)

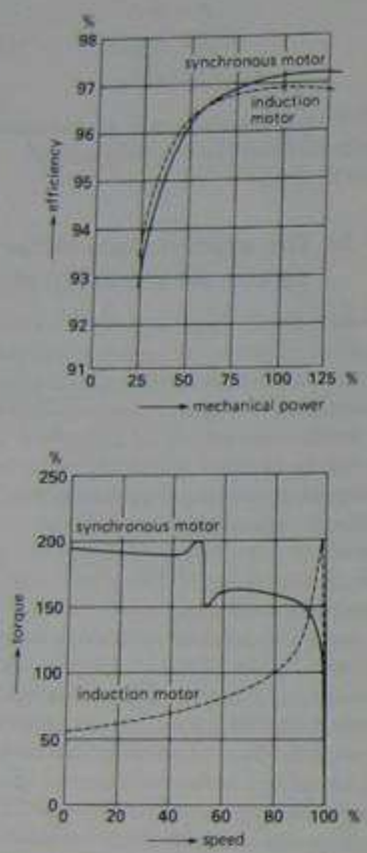


Figure 17.23
Comparison between the efficiency (a) and starting torque (b) of a squirrel-cage induction motor and a synchronous motor, both rated at 4000 hp, 1800 r/min, 6.9 kV, 60 Hz.

inductor) whose reactive power can be varied by changing the dc excitation.

Most synchronous capacitors have ratings that range from 20 Mvar to 200 Mvar and many are hydrogen-cooled (Fig. 17.24). They are started up like synchronous motors. However, if the system cannot furnish the required starting power, a pony motor is used to bring them up to synchronous speed. For example, in one installation, a 160 Mvar

Figure 17.24a
Three-phase, 16 kV, 900 r/min synchronous capacitor rated −200 Mvar (supplying reactive power) to +300 Mvar (absorbing reactive power). It is used to regulate the voltage of a 735 kV transmission line. Other characteristics: mass of rotor: 143 t; rotor diameter: 2670 mm; axial length of stator iron: 3200 mm; air gap length: 39.7 mm.

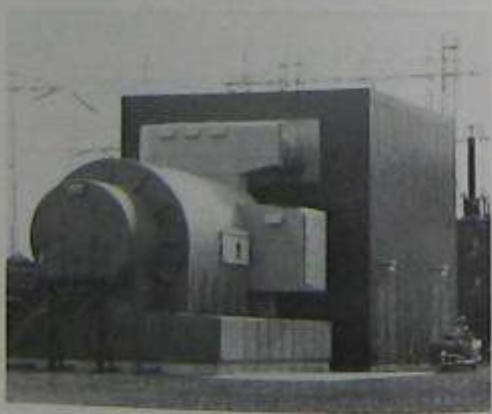


Figure 17.24b
Synchronous capacitor enclosed in its steel housing containing hydrogen under pressure (300 kPa, or about 44 lbf/in^2).
(Courtesy of Hydro-Québec)

synchronous capacitor is started and brought up to speed by means of a 1270 kW wound-rotor motor.

Example 17-7

A synchronous capacitor is rated at 160 Mvar, 16 kV, 1200 r/min, 60 Hz. It has a synchronous reactance of 0.8 pu and is connected to a 16 kV line. Calculate the value of E_o so that the machine

a. Absorbs 160 Mvar
b. Delivers 120 Mvar

Solution

a. The nominal impedance of the machine is

$$Z_n = E_n^2/S_n \qquad (16.3)$$
$$= 16\,000^2/(160 \times 10^6)$$
$$= 1.6\ \Omega$$

The synchronous reactance per phase is

$$X_s = X_S(\text{pu})\, Z_n = 0.8 \times 1.6$$
$$= 1.28\ \Omega$$

The line current for a reactive load of 160 Mvar is

$$I_n = S_n/(\sqrt{3}\, E_n)$$
$$= 160 \times 10^6/(1.73 \times 16\,000)$$
$$= 5780\ \text{A}$$

The drop across the synchronous reactance is

$$E_x = IX_s = 5780 \times 1.28$$
$$= 7400\ \text{V}$$

The line-to-neutral voltage is

$$E = E_L/\sqrt{3} = 16\,000/1.73$$
$$= 9250\ \text{V}$$

Selecting E as the reference phasor, we have

$$E = 9250\angle 0°$$

The current I lags 90° behind E because the machine is absorbing reactive power; consequently,

$$I = 5780\angle -90°$$

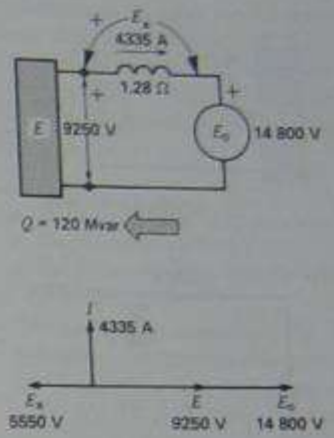


Figure 17.25b
Over-excited synchronous capacitor delivers reactive power (Example 17-7).

From Fig. 17.25a we can write

$$-E + jIX_s + E_o = 0$$

hence

$$E_o = E - jIX_s$$
$$= 9250\angle 0° - 5780 \times 1.28\angle(90° - 90°)$$
$$= 1850\angle 0°$$

Note that the excitation voltage (1850 V) is much less than the line voltage (9250 V).

b. The load current when the machine is delivering 120 Mvar is

$$I_n = Q/(\sqrt{3}\, E_n)$$
$$= 120 \times 10^6/(1.73 \times 16\,000)$$
$$= 4335\ \text{A}$$

This time I leads E by 90° and so

$$I = 4335\angle 90°$$

From Fig. 17.25b we can write

$$E_o = E - jIX_s$$
$$= 9250\angle 0° - 4335 \times 1.28\angle 180°$$
$$= (9250 + 5550)\angle 0°$$
$$= 14\,800\angle 0°$$

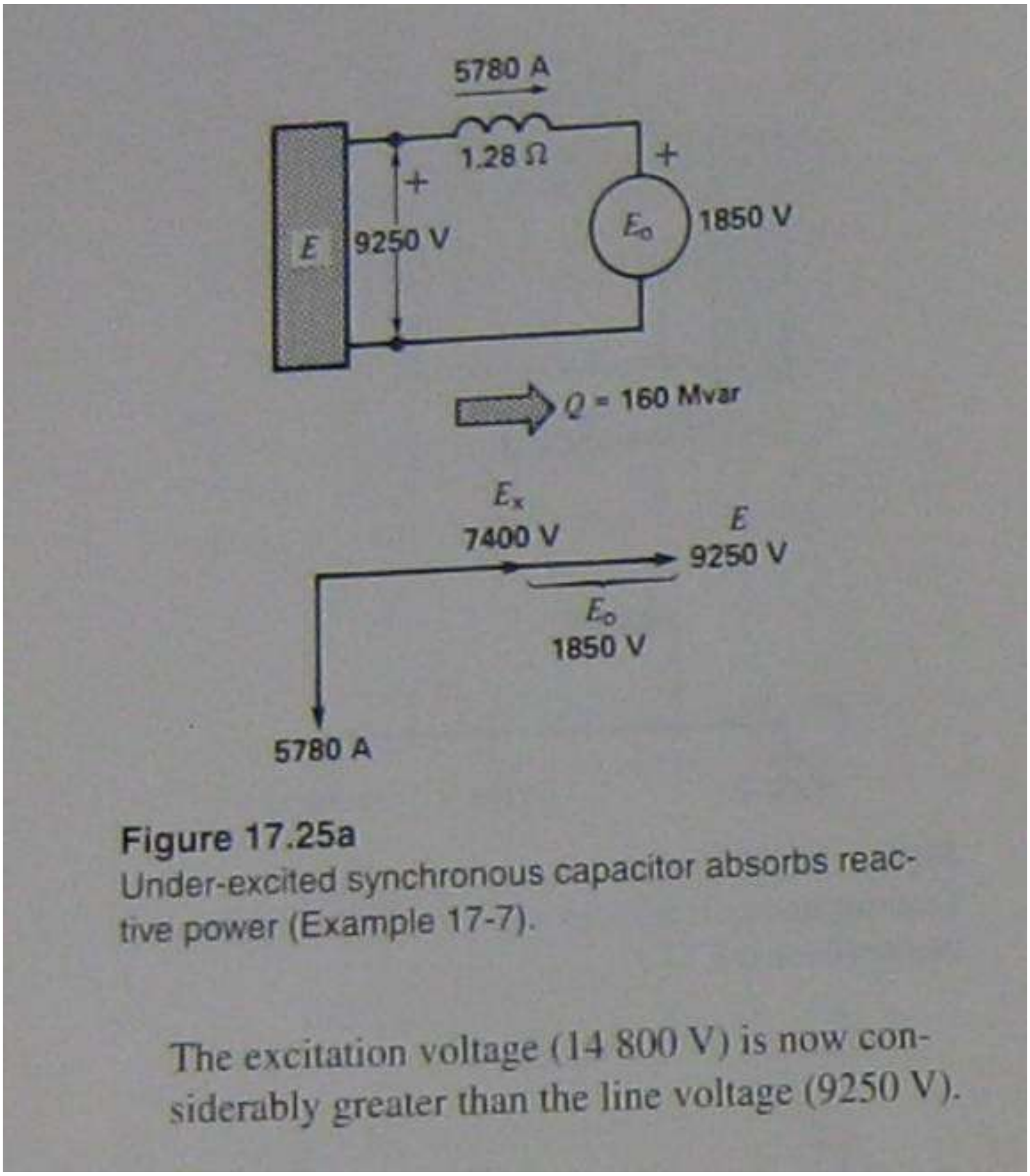


Figure 17.25a
Under-excited synchronous capacitor absorbs reactive power (Example 17-7).

The excitation voltage (14 800 V) is now considerably greater than the line voltage (9250 V).

Learning Outcome 4.6

Describe the causes of hunting in a synchronous motor and the limits for stable operation

EXAMPLE 12.5 A 2 MVA 3-phase 8-pole alternator is connected to 6,000 V 50 Hz busbars and has a synchronous reactance of 4Ω per phase. Calculate the synchronizing power and synchronizing torque per mechanical degree of rotor displacement at no-load. (Assume normal excitation.)

The synchronizing power coefficient is

$$P_s = \frac{3VE_F}{X_s} \cos \sigma \tag{12.44}$$

On no-load the load angle $\sigma = 0$.

Since there are 4 pole-pairs, 1 mechanical degree of displacement is equivalent to 4 electrical degrees; therefore

$$P_s = 3 \times \frac{6{,}000}{\sqrt{3}} \times \frac{6{,}000}{\sqrt{3} \times 4} \times \frac{4}{1{,}000} \times \frac{\pi}{180} = \underline{\underline{627\,\text{kW/mech. deg}}}$$

Synchronous speed of alternator, $n_0 = \dfrac{f}{p} = 12{\cdot}5$ rev/s

Thus

$$2\pi n_0 T_s = 627 \times 10^3$$

and

Synchronizing torque, $T_s = \underline{\underline{8{,}000\,\text{N-m/mech. deg}}}$

12.17 Oscillation of Synchronous Machines

In the previous sections, transient accelerations or decelerations of an alternator rotor were assumed in order to investigate the synchronizing power and synchronizing torque. Such transients may be caused by irregularities in the driving torque of the prime mover or, in the case of a motor, by irregularities in the load torque, or by irregularities in other machines connected in parallel, or by sudden changes in load.

Normally the inherent stability of alternators when running in parallel quickly restores the steady-state condition, but if the effect is sufficiently marked, the machine may break from synchronism. Moreover, if the disturbance is cyclic in effect, recurring at regular intervals, it will produce forced oscillations in the machine rotor. If the frequency of this cyclic disturbance approaches the value of the natural frequency of the rotor, when connected to the busbar system, the rotor may be subject to continuous oscillation and may eventually break from synchronism. This continuous oscillation of the rotor (periods of acceleration and deceleration) is sometimes known as *phase swinging* or *hunting*.

Fig. 12.18 shows the torque/load-angle characteristic of a synchronous generator. The steady input torque is T_0, corresponding to a steady-state load angle σ_0. Suppose a transient disturbance occurs such as to make the rotor depart from the steady state by σ'. Let σ' be sufficiently small to assume that the synchronizing

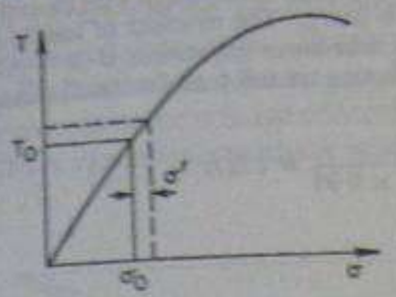


Fig. 12.18 OSCILLATION OF A SYNCHRONOUS MACHINE CONNECTED TO INFINITE BUSBARS

torque is constant; i.e. the torque/load-angle characteristic is assumed to be linear over the range of σ' considered.

Let T_s = Synchronizing torque coefficient (*N-m*/mech. rad)
σ' = Load angle deviation from steady-state position (mech. rad)
J = Moment of inertia of rotating system ($kg\text{-}m^2$)

Assuming that there is no damping,

$$J\frac{d^2\sigma'}{dt^2} = -T_s\sigma' \tag{12.46}$$

The solution of this differential equation is

$$\sigma' = \sigma_m' \sin\left(\sqrt{\frac{T_s}{J}}\,t + \psi\right) \tag{12.47}$$

From eqn. (12.47), the frequency of undamped oscillation is

$$f = \frac{1}{2\pi}\sqrt{\frac{T_s}{J}} \tag{12.48}$$

Synchronous machines intended for operation on infinite busbars are provided with damping windings in order to prevent the sustained oscillations predicted by eqn. (12.48).

In salient-pole machines the damping winding takes the form of a short-circuited cage consisting of copper bars of relatively large cross-section embedded in the rotor pole-face. In cylindrical-rotor machines the solid rotor provides considerable damping, but a cage winding may also be provided. This consists of copper fingers inserted in the rotor slots below the slot wedges and joined together

11.3.5 Hunting in alternators

The driving torque and speed of a piston engine is not absolutely constant during a complete revolution but varies according to the position and speed of the pistons. This causes minute variations in the speed of the alternator shaft. The speed variations are small but cause momentary increases and decreases above and below the average rotational speed. The effect is called *hunting* and leads to small voltage variations, which can include harmonics distorting the waveform.

It is partially neutralised by the inertia of the rotating parts. Remedies for hunting involve the use of quite heavy flywheels and special windings in the pole faces. Called *amortisseur* windings, they are discussed in more detail in section 11.5.5 later in this chapter.

The voltage pulses created by hunting can cause circulating currents to flow between alternators connected in parallel, resulting in an increase in the mechanical oscillations of the rotating parts. Electrical losses are also increased.

High-speed turbines are not affected to the same extent by hunting. Their major cause of oscillation about a fixed point is the minor adjustments of the governors as load changes on the machine occur.

11.4 STANDBY POWER SUPPLIES

Standby power supplies are generally intended to provide mains power at a specified voltage and frequency. There are two main forms of standby power-supply units.

The first type of unit is meant for use where no interruption to a power supply can be tolerated, for example, to computer, hospital, and aircraft navigation equipment. Losing power at a crucial moment in an operation could mean loss of life, or in the middle of a computer operation could mean the loss of valuable data. There is also an increasing use of this type of power supply for portable work because it can often be run from a 12 V vehicle battery. It is quick, convenient and quiet. Built into the vehicle, it is always ready for use.

The second type of standby unit is where momentary losses of power can be tolerated. Such uses would include emergency lighting, theatres, and industrial uses such as fully environmental meat-bird sheds. Delays of several seconds in restoring power can be acceptable in some circumstances. This type of standby power supply would also be suitable for lifts and high-rise buildings.

A subsection of this latter category includes portable power supplies such as small generating plants that can be carried from job to job in a vehicle.

11.4.1 Uninterruptible power supplies (UPS)

A block diagram of a UPS is shown in Figure 11.12. It [...] be seen that the unit runs off the mains supply wi[...] battery permanently 'floating' on charge from an in[...] battery charger. Effectively the battery bank is suppl[...] an inverter, which converts direct current to alterna[...] current at mains voltage and frequency. In the even[...] losing mains power, the unit continues to operate as l[...] as the battery has sufficient charge.

More critical loads usually have an engine-driven al[...] nator on standby to ensure that the battery charg[...] maintained. The battery capacity has to be great eno[...] to supply the circuit power while the alternator [...] engine are being started and run up to speed. Allowa[...] also have to be made for non-starting incidents [...] provide greater flexibility as a backup in an emergen[...]

Direct current values are high, so when using a veh[...] battery, care must be taken to ensure that the battery [...] not go flat. For example, a 500 W television set dr[...] about 20 A on 32 V d.c. The current drain would ex[...] exceed 50 A on 12 V.

More modern inverter units are smaller and may d[...] less current, but they usually have a time rating of a[...] 15 minutes or so and must then be switched off to [...] down. A 500 VA modern unit probably has a full-t[...] rating of about 150 VA.

11.4.2 Engine-driven alternators

A large range of engine-driven alternators is availabl[...] a choice has to be made on the basis of several fac[...] They range from buying a small portable unit at the [...] possible price, to careful planning for the most suit[...] unit for a particular purpose. It is not enough to selec[...] alternator with respect to the load it has to supply[...] choice should take into account many other consid[...] tions. Some of these factors are listed below and t[...] order of importance is governed by the actual inten[...] use for the alternator.

Purchase price

The overall cost of smaller units may be lower, bu[...] terms of cost per kVA they are more expensive and ope[...] at lower efficiencies. As the size of the unit increases, [...] cost per kVA reduces, while the operating effici[...] increases.

Type of prime mover

The economy of the prime mover in terms of efficiency [...] a bearing on its selection. This in turn is affected by [...]

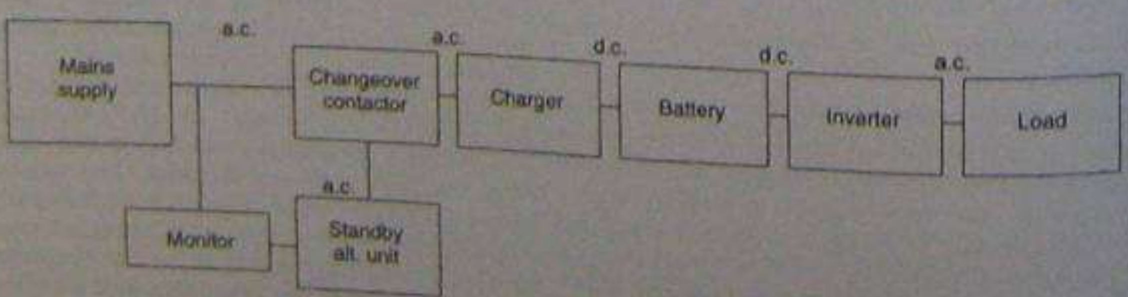


Figure 11.12 • Block diagram for an uninterruptible power supply with backup

type of service it will encounter. For example, a steam turbine has good economy throughout its entire load range. However, it is expensive, large, and needs a long time to get the unit on load from cold. An internal combustion engine has poor efficiency at light loads but is much cheaper to buy initially. For some loads it is cheaper to buy several smaller alternators than one large unit. Problems of paralleling the units then have to be considered (see section 11.3).

The cost and availability of fuel must always be a consideration. While distillate is more expensive initially, as is the diesel engine itself, the fuel cost per hour is less, while maintenance costs are far higher than those for a petrol engine. The petrol engine is cheaper to buy, the fuel is readily available, and the unit is suited to smaller units used purely for portable power supplies on intermittent duties. In the long term the diesel engine runs better on full loads than the petrol engine. The petrol engine is more tolerant of dirty fuel than the diesel engine and does not need specialised skills for maintenance purposes. Figure 11.13 shows a portable generating unit driven by a single-cylinder petrol engine. Two 15 A outlets are available for appliances.

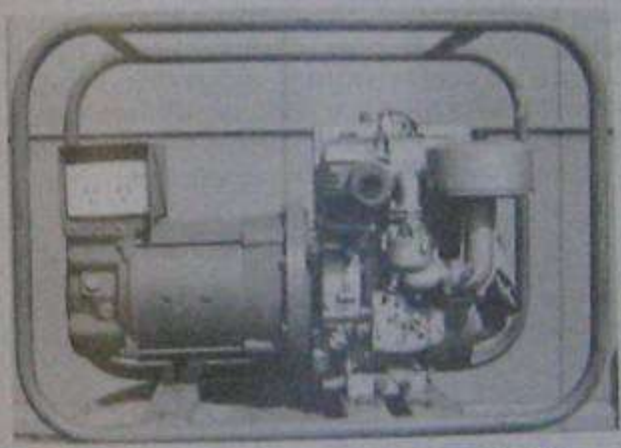


Figure 11.13 • Self-contained portable power supply. The alternator rated at 6 kVA is driven by a petrol engine. The size and weight of the unit is such that it can be carried to any site where power is required

Starting methods

Starting methods are governed by the intended use of the generating unit. The quicker the changeover to auxiliary power, the more expensive is the starting method. The cheapest method involves merely starting the unit manually when it is realised that the main power supply has failed. A more expensive method involves the use of a changeover contactor that drops out when the main supply fails. In turn this connects a starting motor to the engine and when the alternator is up to speed, connects it to the load.

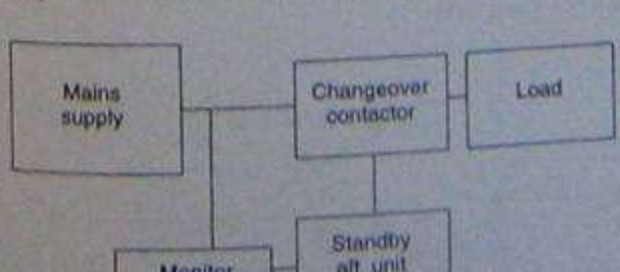


Figure 11.14 • Block diagram for an engine-driven standby alternator

Load sizes and alternator capacities

Smaller generating plant is usually intended for standby purposes for short periods. It usually has only one load connected to it at a time, such as a portable tool or a small lighting load. With middle- and larger-size alternators, consideration has to be given to the possible connection of intermittent larger loads, such as the starting currents of motors. The unit then has to have the electrical capacity and engine power to maintain both the output voltage and the frequency during these current surges to avoid interruptions to other equipment connected to the same supply.

Operation of alternators

With the exception of some manually operated equipment, most operations are now beyond the control of the operator. Where some degree of manipulation is available there are two important factors that should always be considered—voltage and frequency. In most cases the voltage is governed by automatic voltage regulators, while the frequency is controlled by the engine governor. The order of operation is to set the speed first, which in turn sets the frequency, and then adjust the voltage of the unit. To do this in the reverse order is to alter the voltage each time the speed is altered.

11.5 THREE-PHASE SYNCHRONOUS MOTORS

A three-phase synchronous motor has no starting torque. It has to be manipulated up to speed or as close to it as possible so that it can pull itself into synchronism.

Once up to speed, the rotor field can be excited with direct current and the rotor is in effect then dragged around at the same speed as the three-phase stator field. Its speed is synchronised with that of the stator field. This is markedly different in principle to the induction motor, where the rotating field of the stator is pushing against the induced rotor field. That causes the rotor to rotate, but with some slip, whereas in the synchronous motor there cannot be slip, merely a 'hanging back' due to the load imposed on the machine. This is illustrated in Figure 11.15 and shows as a torque angle. If the load becomes too great for a synchronous motor it immediately pulls out of synchronism.

11.5.1 Construction

Stator

The stator has a three-phase winding and is of the same type as that in an alternator or induction motor.

When this winding is energised with a.c. it produces a magnetic flux that rotates at a speed called the synchronous speed. It is the same speed at which the synchronous motor would have to be driven to generate an a.c. voltage at line frequency.

The speed can be derived from the same formula used for alternators in section 11.2.3.

Rotor

Although of similar construction to the alternator rotor, it is usually made with salient poles. When excited with d.c. it produces alternate north and south magnetic poles, which are attracted to those produced in the stator.

11.5.2 Operating principle

A synchronous motor works on the principle of magnetic attraction between two magnetic fields of opposite polarity; one field is that of the rotating stator and the other that of the rotor.

A synchronous motor has torque only at synchronous speed, so special steps have to be taken to get the motor up to speed and synchronised with the supply. The two magnetic fields are then rotating at the same speed and lock in with each other.

11.5.3 Effect of load on a synchronous motor

When a synchronous motor runs on no load, the relative positions of stator and rotor poles coincide as shown in Figure 11.15(a).

When a load is applied, the rotor must still continue to rotate at synchronous speed but owing to the retarding action of the load, the rotor pole lags behind the stator pole. Their relative positions are displaced by (called the 'torque' or 'load' angle), as show 11.15(b). The greater the load applied, the g torque angle.

The magnetic coupling between each stat pole distorts according to the load applied. If the motor becomes excessive, the magne breaks and the rotor slows down until it stops

When the motor is rotating at synchronous a fixed d.c. excitation in the rotor windings, t cuts the stator windings, inducing a voltage in winding and opposing the applied voltage The phase relationship between this induced the applied voltage depends on the relative each stator and rotor pole, which in turn de load applied to the motor.

Neglecting motor losses, on no load the tor zero, and so the induced voltage V_g and voltage V are equal and opposite. The resultan across the windings is zero, and so the current the supply is also zero. This is illustrated by th Figure 11.16(a).

When a light load is applied to the moto angle α increases, and the induced voltag stator windings is now $(180 - \alpha)$°E out of ph applied voltage V, as shown in Figure 11.16(voltages combine to produce an effective across the stator windings, which is sufficie current I from the supply. Because of the re

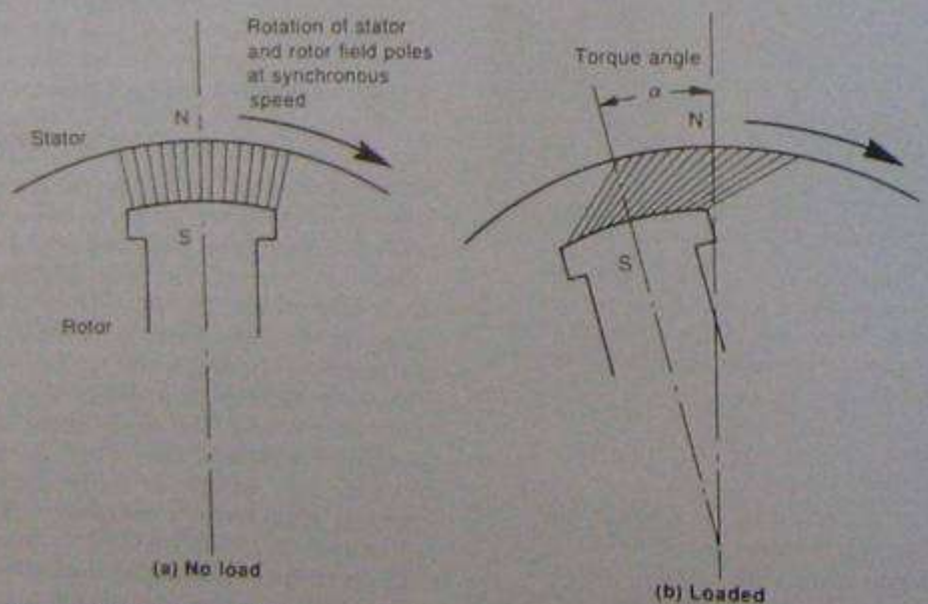


Figure 11.15 • Relative positions of stator and rotor magnetic fields in a synchronous motor

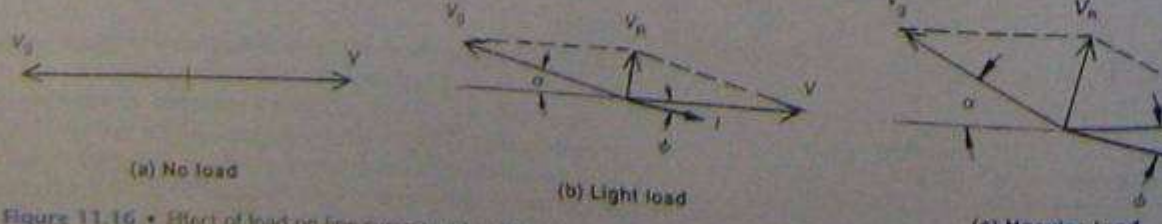


Figure 11.16 • Effect of load on line current with constant excitation

inductance of the stator windings, the line current I in each winding lags each resultant voltage V_R by nearly 90°E. This causes the line current I to lag the applied voltage by ϕ.

As the load is increased, so the torque angle is increased. This causes an increase in the resultant voltage V_R across each stator winding, as seen in Figure 11.16(c). Because of the increase in the value of V_R, the line current I increases, and the phase angle ϕ between the applied voltage V and the line current I also increases.

For fixed excitation, any increase in load on a synchronous motor will cause an increase in current drawn, at a lower power factor.

11.5.4 Effect of varying field excitation

If the load applied to a synchronous motor is constant, the power input to the motor is also constant.

When the rotor field excitation is varied, the induced voltage in each stator winding is also altered.

The phasor diagram in Figure 11.17(a) represents the conditions for a given load at unity power factor. The power input per phase is VI_1. If the rotor field excitation is decreased, the induced voltage V_g decreases, as shown in Figure 11.17(b). This causes the line current I_2 to lag the applied voltage V by ϕ_2. Since the load, and so the power input, is constant, the power component of I_2 must remain the same as I_1 in Figure 11.17(a). The line current I_2 must increase to accommodate the lagging power factor. A reduction in the d.c. field excitation therefore causes an increase in line current, and a lagging power factor.

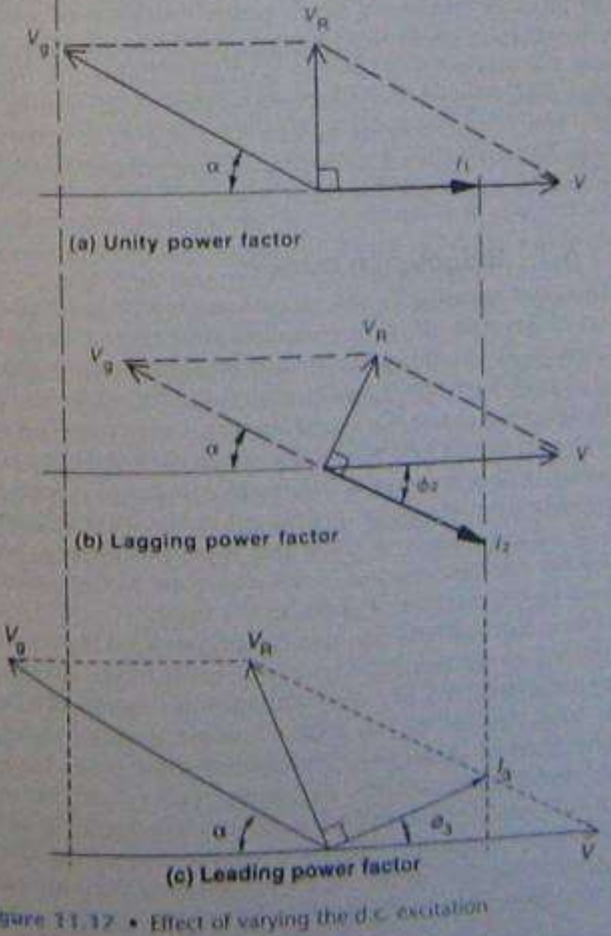


Figure 11.17 • Effect of varying the d.c. excitation

If the d.c. excitation is increased, the induced voltage V_g increases, as shown in Figure 11.17(c). The line current I_3 will therefore lead the applied voltage V by ϕ_3, and will also be greater than I_1 in Figure 11.17(a) because the power component is the same, owing to the load remaining constant. An increase in d.c. excitation therefore causes an increase in line current, and a leading power factor.

It can be seen that if the excitation of a synchronous motor on a constant load is varied from a low to a higher value then:

1. stator current gradually decreases, reaches a minimum, and then increases again
2. the power factor, at first lagging, gradually increases, becomes unity when the stator current is a minimum, and then decreases again, but becomes leading.

Care should be taken when adjusting the excitation of a synchronous motor. There are limits to which it can be taken with safety. Over-excitation and under-excitation can cause the synchronous motor to become unstable. Once these limits have been exceeded, the power produced by the motor decreases and the danger of overloading becomes imminent as the machine exceeds its design limits. The most obvious situation is one of under-excitation where the magnetic bond between the rotating field and the rotor is so weakened that the load exceeds the pull-out torque of the motor and it drops out of synchronism. Over-excitation creates a situation where the line current and mechanical load exceed the full-load rating of the machine and the magnetic bond becomes so stiff that changes in load place undue mechanical stresses on the motor shaft.

11.5.5 Hunting in synchronous motors

A change in load on a synchronous motor causes a change in the value of the torque angle (Fig. 11.15). In general, the inertia of the rotor prevents an instant change to the new conditions, with the result that the rotor shifts past the point of equilibrium and then has to correct itself. While the rotor and the rotating field in the stator are still rotating at a synchronous average speed, the change in load on the rotor causes this periodic swing around the point of equilibrium. This surging or hunting causes an undesirable fluctuation in line current to the motor.

The usual method for damping these surges is to use a damper winding, called an *amortisseur* winding. It consists of copper bars imbedded in the pole faces of the rotor and shorted out at each end (Fig. 11.18). Any surging causes an induced voltage in the copper bars. This results in a magnetic field being created and opposing the surging effect.

Often the shorting-out bars are extended around the rotor, resulting in a squirrel-cage-type rotor winding about the salient poles. While damping any tendency of the rotor to hunt, they can also assist the motor in starting by acting as sections of a squirrel-cage winding. In effect this winding enables the motor to be started as an induction motor.