Electric Drives

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ELECTRIC DRIVES
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PREFACE

The purpose of this book is to provide a thorough grounding in fundamental aspects of electric drives and is, therefore, suitable for use as a text by students pursuing undergraduate courses in electrical engineering and those preparing for AMIE examinations. The principal topics covered in sufficient depth are: motor characteristics, speed control of motors, selection of drive motors, transients in drive systems, motor starters and controllers, and application of drives in industries.

The introductory discussion of Chapter 1 provides an idea of the coverage of the book and outlines important issues involved in design and application of electric drive systems. Chapter 2 discusses the speed-torque characteristics of both dc and ac motors and the methods of braking applied to electric drives. Chapters 3 and 4 deal with the speed control methods of dc and ac motors, respectively. Both the conventional and state-of-the-art methods of speed control, with an example of a microprocessor-based dc controller, are discussed in detail. Several forms of power electronic controllers are also described. Chapters 5 and 6 cover several other aspects of drive systems such as selection of drive motors for different duty cycles and transient behaviour and dynamics of drive systems during starting/braking, etc. Chapter 7 gives a description of various types of controller circuits used in starting/braking of motors. Lastly, Chapter 8 is devoted to a discussion of drive systems as used in different industries such as steel mills, cement mills, electric traction, coal mines, paper mills, machine tools, and textile mills. Each application is covered in depth and supported by appropriate illustrations.

A large number of solved examples and unsolved problems are included in Chapters 2 to 6, necessary to enhance the understanding of the material presented. Answers to all the problems are provided at the end of the book.

We wish to express our gratitude to our teachers who initiated us in this area, and several other colleagues in our respective institutes who helped us in improving the quality of the book. We are also grateful to all of those who assisted in preparing the manuscript.

Finally, we want to express our love and appreciation to our families for their active support, patience and encouragement throughout the time we were writing this book.

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CHAPTER 1

INTRODUCTION

1.1 CONCEPT OF ELECTRIC DRIVES

An electric drive can be defined as an electromechanical device for converting electrical energy into mechanical energy to impart motion to different machines and mechanisms for various kinds of process control.

Electric drives have numerous applications in diverse areas such as industry, transport, agriculture and domestic. Various functions performed by electric drives include the following:

1. driving fans, ventilators, compressors and pumps, etc.,
2. lifting goods by hoists and cranes,
3. imparting motion to conveyors in factories, mines and warehouses, and
4. running excavators and escalators, electric locomotives, trains, cars, trolley buses, lifts and drum winders, etc.

Steel and plastic rolling mills, textile and paper mills, printing presses and machine tools, etc., all incorporate electric drives. Cutting, milling, drilling, grinding, punching, pressing, clipping and other operations of machine tools and industrial installations are all performed by electric drives. Electric drives are also incorporated in flight and landing control systems of aircraft and in propulsion systems of submarines for cruising underwater. It is no exaggeration to say that today the applications of electric drives are as broad and vast as industrialization itself. They provide power outputs from several tenths of watt in recording and controlling instruments to several thousand kilowatts in electric locomotives and rolling mills. The speed ranges from creeping to several thousands of rpm.

The operations of electric drives are selectively regulated and made partly or fully automatic to increase the productivity and efficiency of the industry. The controllers are defined under three heads: (a) Manual, (b) Semiautomatic, and (c) Automatic.

The mode of operation and economics dictate the type of control. The load and its characteristics are the basic parameters for determining the motor rating. The load may be of continuous or intermittent type. The magnitude of the load can be constant or variable. Environmental conditions should always be taken into account. External temperature, cooling method (natural or forced), humidity, density of dust and dirt, vibrations and shocks are also the important factors that determine the design and construction of the mechanical transmission and the type of motor. Electric motors may be dc or ac motors. The types of control gear used may be electromechanical, electronic (solid state), and so on.

1.2 CLASSIFICATION OF ELECTRIC DRIVES

According to developments that have taken place (Table 1.1), electric drives in industrial applications may be classified into three types: group, individual and multi-motor drive systems.

Table 1.1 Classification of Electric Drive Systems

Loads and environment	Loads	Continuous or intermittent, positive, negative or varying sign, constant or variable in magnitude
	Environment	Temperature, humidity, dust and dirt, vibrations and shocks
Mode of operation	Continuous duty	
	Short time duty	
	Intermittent duty	
Means of control	Manual	
	Semiautomatic	
	Automatic	
Number of machines and interrelations	Individual drive	
	Group drive	
	Multi-motor drive	Independent, non-related control
		Interdependent, related control
Dynamic and transients	Uncontrolled transient period	
	Controlled transient period	
Methods of speed control	Reversible and nonreversible uncontrolled constant speed	
	Reversible and nonreversible step speed control	
	Reversible and nonreversible smooth speed control	
	Constant predetermined position control	
	Variable position control (servomechanism)	
	Composite control	

1.2.1 Group Electric Drive

This drive consists of a single motor, which drives one or more line shafts supported on bearings. The line shafts may be fitted with either pulleys and belts or gears, by means of which a group of machines or mechanisms may be operated. It is also sometimes called the shaft drive. Group drives are seldom used now due to their low efficiency and other disadvantages, but such drives are more attractive economically.

Advantages

1. A single large motor can be used instead of a number of small motors.

2. The rating of the single large motor may be appropriately reduced taking into account the diversity factor of loads; the motor, normally induction type, can thus work at about full load, increasing the efficiency and power factor.

Disadvantages

1. There is no flexibility. If the single motor used develops fault, the whole process will come to stop.

2. Addition of an extra machine to the main shaft is difficult.

3. If some of the machines are not working, the losses are increased, thus decreasing the efficiency and power factor.

1.2.2 Individual Electric Drive

In this drive, each individual machine is driven by a separate motor. This motor also imparts motion to various other parts of the machine. Examples of such machines are single spindle drilling machines (universal motor is used) and lathes. In a lathe, the motor rotates the spindle, moves the feed and also with the help of gears, transmits motion to lubricating and cooling pumps. A three-phase squirrel cage induction motor is used as the drive. In many such applications, the electric motor forms an integral part of the machine.

1.2.3 Multi-motor Electric Drive

In this drive system, there are several drives, each of which serves to actuate one of the working parts of the driven mechanism. Applications of such a drive are found in complicated metal cutting machine tools, paper making machines, rolling mills, etc. The drives of a crane can also be considered as an example of a multi-motor drive system. This type of multi-motor drive incorporates three drives: first for vertical movement, second for side movement, and third for forward movement of the load. Each of these drives functions separately and the operator of the crane coordinates their functions. But there are some cases, when an automatic coordinated functioning of several drives is necessary. For instance, a traction drive of an electric locomotive consists of 4–8 motors. It is obvious that functioning of these motors should be coordinated in such a way that none loads the other.

A rolling mill incorporates several individual drives. To eliminate tearing and breakage of the product (rods, wires, plates), the speeds of individual drives should be coordinated to maintain continuous uninterrupted flow of material at uniform speed and constant tension. Paper mill drives and cloth printing machines are other examples of multi-motor drives with coordinated functioning.

1.3 CLASSIFICATION OF CONTROL SCHEMES

From the point of view of the means, and consequently the ways of controlling the functioning of drives, three types of control schemes are identified (Table 1.1), namely manual, semiautomatic and automatic.

1.3.1 Manual Control

The electric drives with manual control can be as simple as a room fan, incorporating one switch and a resistance for setting the required speed. It can be more complicated as in the case of a push button starter for a simple machine tool or a hoist drive, or drum or cam controllers for heavy duty crane drives and electric locomotives.

1.3.2 Semiautomatic Control

The electric drives with semiautomatic control usually consist of a manual device for giving a certain command (starting, braking, reversing, change of speed) and an automatic device that, in response to a command, operates the drive in accordance with a preset sequence or order.

1.3.3 Automatic Control

The electric drives with automatic control have a control gear, without manual devices (though as a security measure, such drives usually have a manual device for use, in case the automatic system fails).

1.4 CLASSIFICATION OF METHODS OF SPEED CONTROL

From the point of view of speed control, the electric drives are divided into five major groups (Table 1.1), although the outputs, modes of operation and means of automatic control of electric drives are all different depending upon the application.

1.4.1 The First Group

These drives provide uncontrolled constant speed of rotation during normal operating conditions. The speed may be unidirectional or otherwise depending upon the requirement of a particular mechanism.

The magnitude of speed is not an important consideration and is permitted to fluctuate in response to variation in the magnitude and direction of the load torque. The drives for pumps, ventilators and compressors may serve as good examples. For bidirectional rotation, the drives of a crane can be cited as examples.

1.4.2 The Second Group

These drives provide step changes in speed during normal operating conditions, while maintaining constant speed within the step. Rotation may be unidirectional or otherwise with or without speed control during transient periods. These drives are used in various machine tools. For example, in a drilling machine, an electric drive has to provide different speeds for different materials of the work-piece and diameters of the drill. Step control of speed is used in heavy duty cranes, lifts and mine winders in order to have appropriate speeds.

1.4.3 The Third Group

These drives provide smooth, or with narrow step, control of speed within certain ranges, during normal operating conditions. A drive of an electric locomotive can serve as an example, where smooth speed control of the traction motor is required. The main drive of a rolling mill can serve as another example of drives of this group, but compared to a locomotive traction drive, the requirements of speed control here are different.

The main drive of a blooming mill must provide minimum decrease of speed, when the billet is passing through the rolls, and a fast reversal. The drives of a continuous rod and plate rolling mill, printing presses must rotate in one direction, but their speeds must be synchronized and accurately controlled so that the products (rods, plates, paper and linolium sheets) are uniformly rolled.

1.4.4 The Fourth Group

These drives have to provide control for fixed positions of the load. Drives for lifts and mine winders are examples of drives of this group. This type of drive is used for automatic production lines, skip lifting and loading installations in sugar and chemical industries.

1.4.5 The Fifth Group

These drives are used for position control in response to a command. They are also called servomechanisms. They are widely used in recording and controlling instruments, copying devices for machine tools, drive systems for radio and radar antennas, etc.

1.5 COMPONENTS OF ELECTRIC DRIVES

With the introduction of Thyristor or Silicon Controlled Rectifier (SCR) as a power switching device in the sixties, and later the arrival of Gate Turn-Off (GTO) thyristor, bridge rectifiers (converters) and choppers fed from ac and dc supply respectively are used for speed control of dc motors. Choppers using GTOs are now preferred for high power dc drives as these devices can be turned off by proper signals applied to the gate. This is unlike thyristors, which can only be turned on by gate signal, i.e. cannot be turned off by gate signal. The inverters, fed from dc supply, are fabricated using GTOs or other devices from transistor family such as Bipolar Junction Transistors (BJTs), Insulated Gate Bipolar Transistors (IGBTs) and Metal Oxide Semiconductor Field Effect Transistors (MOSFETs). These are self-commutated devices, switched on or off by proper base drive, unlike thyristors, turned off only by forced commutation using circuits consisting of inductors and capacitors in conjunction with additional thyristors. Transistor chopper-fed dc drives are used for low and medium power applications. Two types of power electronic converters, namely inverters (mostly voltage source type) and cycloconverters, fabricated normally using thyristors, are used for speed control of ac motors, both induction and synchronous. Both types can provide variable voltage variable frequency (VVVF) control with constant voltage/frequency ratio to produce rated (constant) flux in the air gap of ac motors, at frequencies lower than base (rated) frequency. Sophisticated control strategies such as vector or field oriented control and other control laws developed using modern control theoretic techniques, can now be implemented using microprocessors, microcontrollers, Personal Computers (PCs), and/or Digital Signal Processors (DSPs). The time needed for computation of control inputs has reduced with increase in clock frequency and higher capability for arithmetic and other functions, along with improved reliability due to reduced chip count in processor/PC-based systems.

1.5.1 Motors

The motors are either dc or ac, fed from dc and ac supply respectively. Various types of dc motors, such as separately excited or shunt, series and compound, are used to satisfy the speed and torque requirements of the drive. Normally, dc shunt motors are used for applications, where precise speed control is needed, while series motors are preferred for traction and fan type loads, including compressors.

AC motors are of two types—induction and synchronous. Induction motors are fitted with either squirrel cage or wound (slip ring) rotor. The first type is cheap, rugged and maintenance free. With the availability of inverters with VVVF control, variable speed ac drives using induction motors are emerging as a powerful alternative to dc drives. The use of Pulse Width Modulation (PWM) techniques in inverters has resulted in reduction of harmonics in output voltage waveforms and higher efficiency along with reduction of torque pulsations. In cases where high starting torque is required, the wound rotor (slip ring) motor is preferred, even though the cost is higher. The rotor resistance can be varied, by using a GTO/transistor chopper in parallel with an additional resistance and changing its ON time, to control motor speed. Line commutated (phase controlled) rectifiers/inverters are used in the rotor circuit to run motors at sub-synchronous/super-synchronous speeds with slip power being either fed back or taken from the supply via a step-up transformer. For sub-synchronous speed operation, in which case slip power from the rotor taken via slip rings is fed to the supply, controlled (thyristor) bridge rectifier is changed to uncontrolled (diode) rectifier.

Synchronous motors operate at different power factors including leading, by varying field current (excitation), whereas induction motors take current at lagging power factor due to magnetizing component required to produce flux in the air gap. These motors are used for low speed, high power applications in cement and ball mills. Compared to conventional (wound field) type, variable reluctance (with no field winding) and permanent magnet (assumed as constant excitation) type motors are also coming into prominence for low power applications. Another development is the use of 'self-controlled' type, where the base drive or the triggering signals for the devices in inverters supplying power to the stator, are derived from the rotor position sensors. These motors run at different speeds, unlike the motors fed from variable frequency supply which cannot withstand large load torque variations due to stability problems.

1.5.2 Power Electronic Controllers

A power electronic controller consists of basic circuits termed power electronic converters, and other gate or base drive circuits needed to switch on/off the devices including analog/digital chips, but may not include any processor/PC-based system for control law implementation. The converter circuits are of the following types:

1. Bridge rectifiers (converters) and choppers for dc motors.
2. Inverters and cycloconverters for ac motors.

The controlled rectifiers are fabricated using mostly thyristors, though for low voltage output as needed for low speed operation of dc motors, the power factor is low as delay angle (phase) is large. In certain cases, GTOs or transistors are used to obtain symmetrical output voltage as these devices can be switched off by gate/base drive, not needing additional circuits for turning off like thyristors. This results in near unity power factor. If one controlled rectifier is used for motor operation in forward direction, it cannot be used for braking operation in forward direction as thyristors are unidirectional devices. However, the same rectifier can be used to brake the motor in reverse direction as needed for crane/hoist operation. To brake the motor in forward direction, either a reversing contactor with one rectifier or two rectifiers in back-to-back connection (dual converter) are needed. Such circuits are also suitable for four-quadrant operation of the drive. As stated, GTOs/ transistors are used for chopper-fed dc series motor drives. Both motor and braking operations can be performed, but two choppers are used as current has to be reversed in armature circuit of the dc motor for braking.

Variable frequency inverter (voltage source) circuits using GTOs/transistors have become simpler and occupy less space compared to thyristorized circuits needing bulky commutation circuits. The use of PWM technique in such inverters has resulted in their application to induction motor drives. Cycloconverters using thyristors are used to obtain low frequency needed to run synchronous motors at low speed.

Power electronic converter circuits require gate/drive signals to switch on/off the devices. These signals are obtained using different chips and transistors, which are needed at the last stage to obtain proper drive signals of required power (voltage/current) rating. These circuits perform various operations including sequencing and lock-out functions. Processor/PC-based systems are used to implement complex control algorithms to suit the requirements of dc/ac drive systems for better performance.

Electric drive systems perform numerous operations such as frequent starting and stopping of motors including braking, changing connections as needed, running one or more motors in a particular sequence, interlocking to prevent malfunctioning, and disconnecting the motor from supply when abnormal or faulty conditions are detected. Programmable Logic Controllers (PLCs) are used to reduce the hardware requirements, while at the same time software can be used to make changes in sequencing and other functions, if required, thus making the system flexible.

CHAPTER 2

SPEED-TORQUE CHARACTERISTICS OF MOTORS

2.1 BASIC PARAMETERS

The performance of a motor can be easily estimated from its speed-torque characteristics. For selection of a motor and its control circuits, it is necessary to know the torque to be delivered and the speed at which the machine would have to be run in order to obtain the desired output. It is also possible to predict the performance of a motor, if the motor circuit or loading conditions are changed, as required. Torque may be acting in one direction under a certain situation, but may act in the opposite direction as the situation changes. The direction of rotation may also change accordingly.

For the construction of speed-torque characteristics, it is customary to plot the torque along the x-axis and the speed along the y-axis. Speed-torque characteristics are plotted in all the four quadrants, hence speed and torque may be positive or negative depending on the type of operation. It is, therefore, necessary to establish the convention as regards the sign of speed and torque. For most drives, the positive sense of rotation is taken to be the anticlockwise direction. For reversible drives, the positive sense of rotation is normally chosen to be the same. Similarly, the motor torque is considered to be positive, if it causes an increase in speed in the positive direction.

In industrial applications such as cranes or hoists, it is often necessary to stop the motor quickly and accurately and also to reverse the direction of rotation in a controlled manner, for which braking torque is required. For rolling mill drives, the motor has to accelerate from standstill to rated speed in the forward direction, run at that speed for some time, decelerate to standstill, and then accelerate again to rated speed in the reverse direction. The motor, thus, operates in different quadrants under different operating conditions. The quadrantal diagram and speed-torque characteristics depicting the four-quadrant operation are described below for different types of dc motors—separately excited (or shunt), series, and compound—and induction motors.

2.2 TYPES OF LOADS

A positive load torque is opposite in direction to the positive motor torque. The load torque may remain constant, and be practically independent of speed or vary with some power of speed. Besides load torque, there may be other types of loads as enumerated below:

Load for which torque varies as 'power' of speed. There are some loads for which the torque increases with some power of speed. Such is the case, for example, with fans and centrifugal pumps for which the torque varies as square of the speed (Fig. 2.1).

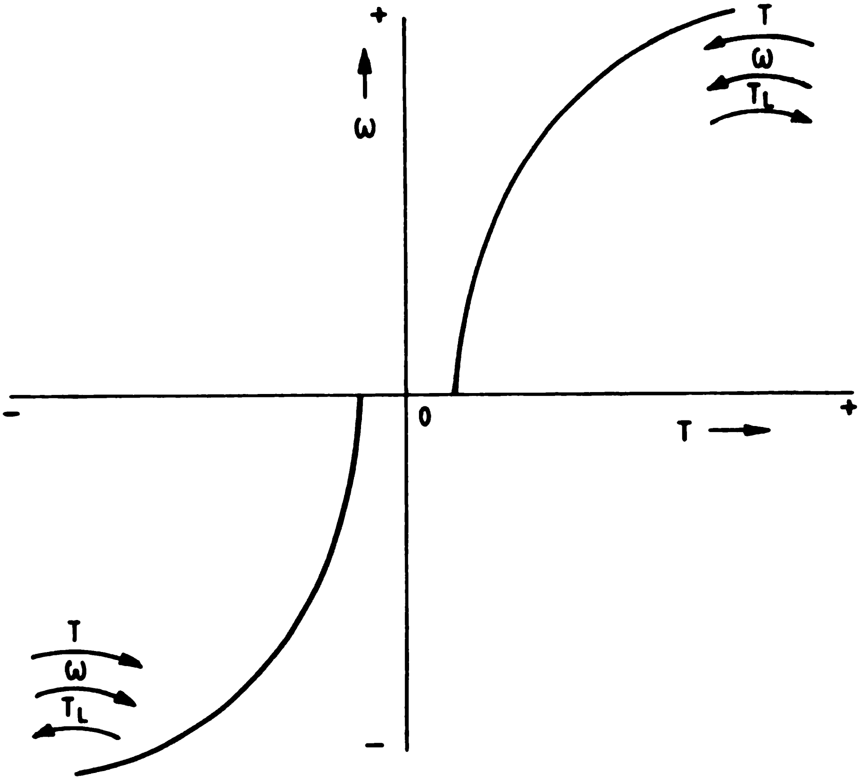


Fig. 2.1 Speed-torque characteristics of a fan load.

Constant friction type of load. Friction torque opposes rotation. It is independent of speed, but passive. It is very small during a normal operation. Sticking friction at the time of starting may be considerable. But almost negligible, if the drive is started frequently. It is also negligible, if the machine is fitted with some form of ball or roller bearings.

Dynamic or inertia load. When the speed of a rotating mass is changing, dynamic torque or inertia torque acts against the motor torque (Fig. 2.2). This torque is given by $J\frac{d\omega}{dt}$, where J is the equivalent inertia of all rotating masses.

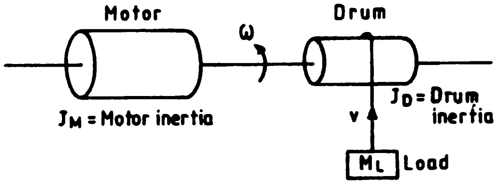


Fig. 2.2 Inertia of dynamic loads.

2.2.1 Quadrantal Diagram

The preceding discussion is illustrated by a quadrantal diagram (Fig. 2.3) showing the four-quadrant operation of a hoist. The load torque is assumed to remain constant and independent of speed. The different modes of operation of a hoist have been depicted. The speed-torque characteristics corresponding to different modes of operation have been illustrated in later sections.

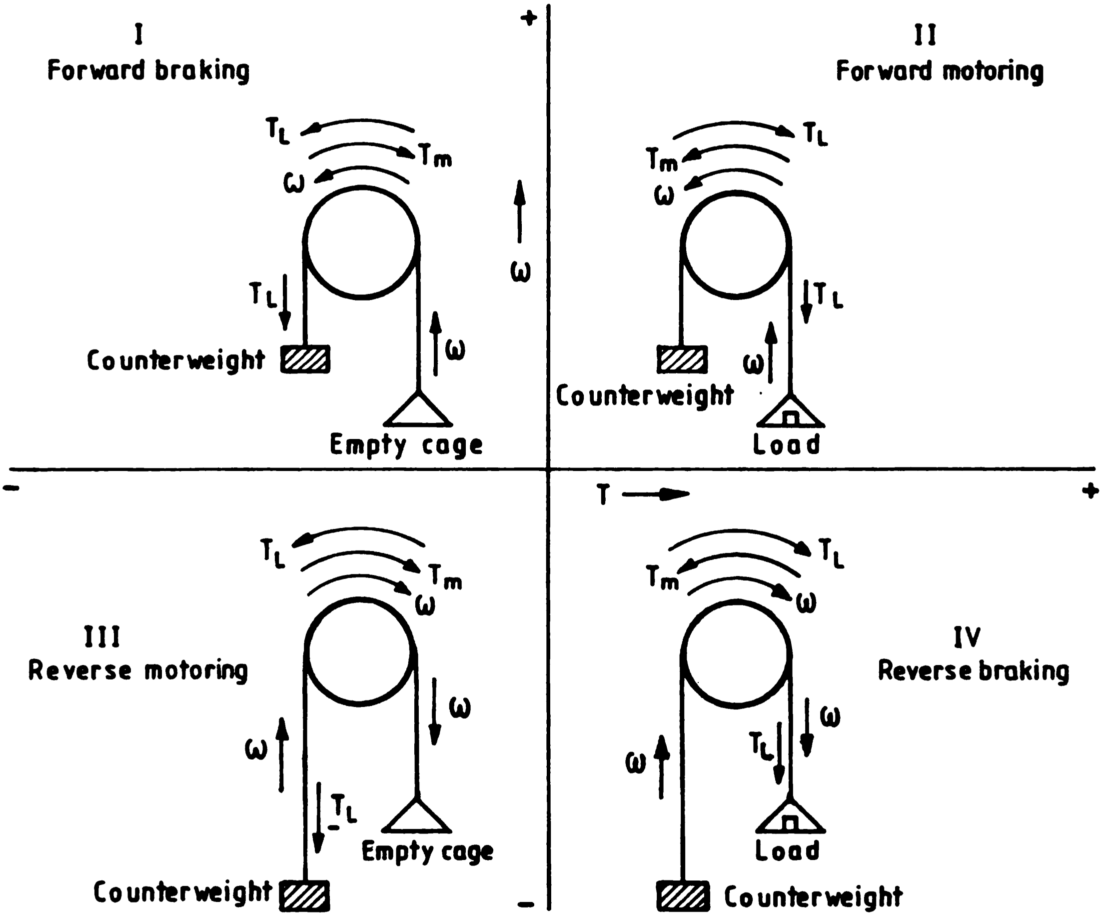


Fig. 2.3 Quadrantal diagram of speed-torque characteristics.

2.3 SPEED-TORQUE CHARACTERISTICS OF DC SHUNT MOTORS

Shunt motors are commonly connected as shown in Fig. 2.4. During steady state operation of the motor, the voltage equation for the armature circuit is given by

$$V = E_b + I_a R \tag{2.1}$$

where

V is the input voltage,

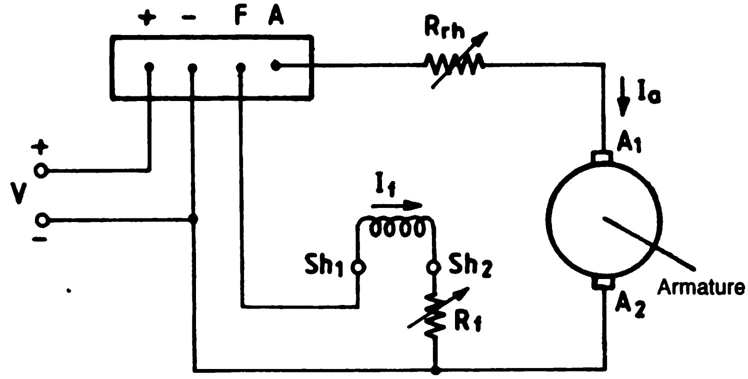


Fig. 2.4 DC shunt motor connections through a four-point starter.

E_b is the back emf,

I_a is the armature current, and

R is the total resistance in the armature circuit given by, $R = R_a + R_{rh}$.

R_a and R_{rh} being respectively the resistance of the armature and the external resistance in series with the armature.

The back emf E_b is expressed by the equation

$$E_b = K\phi\omega \tag{2.2}$$

where

$$K = \frac{Zp}{2\pi a}$$

Z = no. of conductors in the armature,

a = no. of parallel paths in the armature circuit,

p = no. of poles,

ϕ = flux per pole, and

ω = angular speed of the motor in rad/s.

The speed equation is given by

$$\omega = \frac{V - I_a R}{K\phi} \tag{2.3}$$

Torque in N · m is expressed as

$$T = K\phi I_a \tag{2.4}$$

Substituting $\frac{T}{K\phi}$ for the current I_a in Eq. (2.3), the equation for the speed-torque characteristic is obtained as

$$\omega = \frac{V}{K\phi} - \frac{TR}{(K\phi)^2}$$

Putting $C = K\phi$ in the above equation, we have

$$\omega = \frac{V}{C} - \frac{TR}{C^2} \tag{2.5}$$

The coefficient C may be assumed to be constant as field current or flux is kept constant for normal operation in a shunt motor. When V and R are kept constant, the speed-torque characteristic is a straight line. It assumes a drooping shape depending upon the value of the external resistance in the armature circuit. Moreover, when $T = 0$, all characteristics, whatever may be the value of the external resistance in the armature circuit, pass through the same point on the y-axis. This speed is known as the *ideal no load speed* and determined from the equation

$$\omega_0 = \frac{V}{K\phi} = \frac{V}{C} \tag{2.6}$$

The second term in the right hand side of Eq. (2.5), therefore, represents the drop in speed corresponding to the torque and thus

$$\Delta\omega = \frac{TR}{(K\phi)^2} = \frac{TR}{C^2} \tag{2.7}$$

The speed-torque characteristic of the motor without any external resistance in the armature circuit is called the *natural characteristic*. The family of curves for different values of the external resistance in the armature circuit is shown in Fig. 2.5.

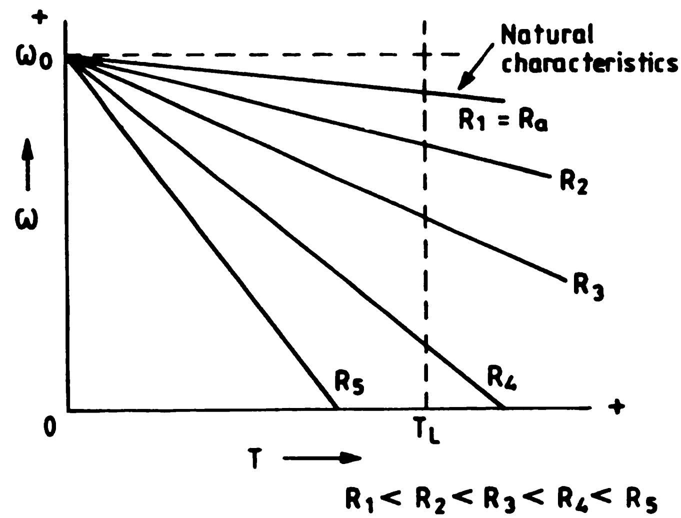


Fig. 2.5 Natural and rheostatic speed-torque characteristics of dc shunt motors.

2.3.1 Construction of Speed-torque Characteristics

From Eq. (2.3), we can write for the natural characteristic

$$K\phi\omega_L = V - I_L R_a$$

where

I_L = full load armature current, and

ω_L = full load speed of the motor.

Thus

$$\omega_L = \frac{V - I_L R_a}{K\phi}$$

and

$$\omega_0 = \frac{V}{K\phi}$$

Therefore, we have

$$\omega_L = \left(1 - \frac{I_L R_a}{V}\right)\omega_0 \tag{2.8}$$

The armature resistance R_a, which is not given in the nameplate data, can be determined approximately by assuming that half of the losses of the motor at rated torque constitute copper loss, i.e. by assuming that maximum efficiency occurs at rated load. Thus, we can write

$$I_L^2 R_a = 0.5(1 - \eta_L)VI_L$$

or

$$R_a = 0.5(1 - \eta_L)\left(\frac{V}{I_L}\right) \tag{2.9}$$

The rated efficiency η_L is taken from the nameplate data. The rated torque is determined from Eq. (2.4)

$$T_L = K\phi I_L = CI_L \tag{2.10}$$

All speed-torque characteristics are straight lines and pass through the point ($T = 0$, $\omega = \omega_0$). To draw a speed-torque characteristic, another point is needed. The natural characteristic of the motor is drawn through the points ($T = 0$, $\omega = \omega_0$) and ($T = T_L$, $\omega = \omega_L$).

The speed of the motor corresponding to rated torque for any value of R_{rh} inserted into the armature circuit is determined from the equation

$$\omega_{rh,L} = \left(1 - \frac{I_L(R_a + R_{rh})}{V}\right)\omega_0 \tag{2.11a}$$

The speed-torque characteristic may also be constructed by using the point of ideal no load speed and the point corresponding to the short circuit of the armature. The short circuit torque T_{sc} is obtained from the equation

$$T_{sc} = \frac{I_{sc}}{I_L}T_L$$

where the short circuit current, $I_{sc} = \dfrac{V}{R}$.

From Eq. (2.5), we have

$$\omega = \frac{V}{C} - \frac{TR}{C^2} = \frac{V}{C}\left(1 - \frac{R}{CV}\cdot T\right)$$

or

$$\omega = \omega_0\left(1 - \frac{T}{T_{sc}}\right) \tag{2.11b}$$

Equation (2.11b) may also be expressed as

$$\omega = \omega_0\left(1 - \frac{IR}{V}\right).$$

Example 2.1

A 220 V, 21 A, 1000 rpm dc shunt motor has armature resistance of 0.05 Ω and field resistance of 220 Ω. The magnetization curve for the machine is given in the following table:

Field current, A	0.2	0.4	0.6	0.8	1.0	1.2	1.4
Emf at 1000 rpm, V	50	100	150	190	219	235	245

Calculate the speed and the torque developed by the motor with full load current under each of the following conditions:

(a) no external resistance is included,

(b) a resistance of 0.05 Ω is connected in series with the armature, and

(c) a resistance of 110 Ω is connected in series with the field winding.

Solution

$V = 220$ V, $R_f = 220$ Ω, $I_i = 21$ A, $R_a = 0.05$ Ω, $N = 1000$ rpm

(a) Field current $I_f = \frac{V}{R_f} = \frac{220}{220} = 1$ A

Armature current (full load) $I_L = I_i - I_f = 21 - 1 = 20$ A

$$\omega = \frac{2\pi N}{60} = \frac{2\pi \times 1000}{60} = 104.72 \text{ rad/s}$$

$$C = K\phi = \frac{V - I_L R_a}{\omega} = \frac{220 - 20\times 0.05}{104.72} = 2.09 \text{ V}\cdot\text{s/rad or N}\cdot\text{m/A}$$

The graph is linear, so only two points are needed.

$$\omega_0 = \frac{V}{C} = \frac{220}{2.09} = 105.26 \text{ rad/s}$$

$$N_0 = \frac{60\omega_0}{2\pi} = \frac{60 \times 105.26}{2\pi} = 1005 \text{ rpm}$$

$$\omega_L = \left(1 - \frac{I_L R_a}{V}\right)\omega_0 = \left(1 - \frac{20 \times 0.05}{220}\right) \times 105.26 = 104.78 \text{ rad/s}$$

$$N_L = \frac{60 \times 104.78}{2\pi} = 1000 \text{ rpm}$$

$$T_L = CI_L = 2.09 \times 20 = 41.8 \text{ N} \cdot \text{m}$$

(b) $R_{rh} = 0.05\ \Omega$, $R = R_a + R_{rh} = 0.05 + 0.05 = 0.1\ \Omega$, $I_L = 20$ A

$$\omega_{rh,\,L} = \left(1 - \frac{I_L R}{V}\right)\omega_0 = \left(1 - \frac{20 \times 0.1}{220}\right) \times 105.26 = 104.3 \text{ rad/s}$$

$$N_{rh,\,L} = \frac{60 \times 104.3}{2\pi} = 996 \text{ rpm}$$

$$T_L = 41.8 \text{ N} \cdot \text{m}$$

(c) A resistance of 110 Ω is inserted into the field circuit.

$$I_f = \frac{V}{R_f + R'} = \frac{220}{220 + 110} = 0.667 \text{ A}$$

From the magnetization curve, $E_b = 162$ V

$$C = \frac{E_b}{\omega} = \frac{162}{104.72} = 1.55 \text{ V} \cdot \text{s/rad or N} \cdot \text{m/A}$$

$$\omega_0 = \frac{V}{C} = \frac{220}{1.55} = 142.0 \text{ rad/s}$$

$$N_0 = \frac{60 \times 142.0}{2\pi} = 1355 \text{ rpm}$$

$$T_L = CI_L = 1.55 \times 20 = 31 \text{ N} \cdot \text{m}$$

$$\omega_L = \left(1 - \frac{I_L R_a}{V}\right)\omega_0 = \left(1 - \frac{20 \times 0.05}{220}\right) \times 142.0 = 141.35 \text{ rad/s}$$

$$N_L = \frac{60 \times 141.35}{2\pi} = 1350 \text{ rpm}$$

Example 2.2

A 400 V, 15 kW, dc shunt motor takes 42 A, and runs at a speed of 1200 rpm. The shunt field resistance is 200 Ω. Assume that the load torque varies as the square of the speed. Neglect iron and friction losses, and also the armature reaction. Calculate

(a) the resistance to be connected in series with the armature to reduce the speed to 1000 rpm. and

(b) the voltage to be applied to the armature for 1000 rpm, had the control been accomplished by Ward-Leonard method.

Solution

$V = 400$ V, $\quad I_i = 42$ A, $\quad P_o = 15$ kW, $\quad R_f = 200\ \Omega$,

$$N_1 = 1200 \text{ rpm}, \qquad \omega_1 = \frac{2\pi N_1}{60} = \frac{2\pi \times 1200}{60} = 125.7 \text{ rad/s}$$

$$\eta = \frac{P_o}{P_i} = \frac{P_o}{VI_i} = \frac{15 \times 10^3}{400 \times 42} = 0.89$$

$$I_f = \frac{V}{R_f} = \frac{400}{200} = 2 \text{ A}$$

(a) Armature current (full load) $I_L = I_i - I_f = 42 - 2 = 40$ A

$$R_a = 0.5(1 - \eta)\left(\frac{V}{I_L}\right) = 0.5(1 - 0.89)\left(\frac{400}{40}\right) = 0.55\ \Omega$$

$$C = \frac{V - I_L R_a}{\omega_1} = \frac{400 - (40 \times 0.55)}{125.7} = 3.0 \text{ V} \cdot \text{s/rad or N} \cdot \text{m/A}$$

$$T_1 = CI_L = 3.0 \times 40.0 = 120.0 \text{ N} \cdot \text{m}$$

$$N_2 = 1000 \text{ rpm}, \qquad \omega_2 = \frac{2\pi \times 1000}{60} = 104.72 \text{ rad/s}$$

Now, $T \propto N^2$

$$T_2 = \left(\frac{N_2}{N_1}\right)^2 T_1 = \left(\frac{1000}{1200}\right)^2 \times 120 = 83.33 \text{ N} \cdot \text{m}$$

$$\omega_2 = \frac{V}{C} - \frac{T_2 R}{C^2}$$

$$104.72 = \frac{400}{3.0} - \frac{R \times 83.33}{(3.0)^2} = 133.33 - 9.26 \times R$$

or

$$R = \frac{28.61}{9.26} = 3.09\ \Omega$$

Therefore, external resistance to be connected in series $R_{rh} = R - R_a = 3.09 - 0.55 = 2.54\ \Omega$

(b) $T_2 = 83.33\ \text{N}\cdot\text{m}$

$$\omega_2 = \frac{V}{C} - \frac{T_2 R_a}{C^2}$$

or

$$104.72 = \frac{V}{3.0} - \frac{83.33 \times 0.55}{(3.0)^2} = \frac{V}{3} - 5.1$$

or

$$V = 329.5\ \text{V}$$

2.4 TYPES OF BRAKING

It is often necessary to brake the motor rapidly and smoothly according to a given speed schedule. Braking torque may be applied by electrical or mechanical methods. Electrical methods are more precise than mechanical methods, when accurate and smooth control of the stopping instant is required. Braking torque is also needed over some part of the duty cycle and for emergency braking in respect of some applications such as cranes. Reversal and speed control of drives may also be undertaken by means of electric braking.

There are three main methods of electric braking, which may be applied to electrical drives:

(a) Dynamic or rheostatic braking,

(b) Plugging or counter current braking, and

(c) Regenerative braking.

2.4.1 Dynamic or Rheostatic Braking

The supply to the field winding is maintained, but the armature is disconnected from the supply voltage and reconnected to an external resistor (Fig. 2.6a). The machine now acts as a generator, converting kinetic energy stored in its moving parts to electrical energy, which is dissipated as heat in the resistor. This method of braking is called the *dynamic* or *rheostatic* braking.

Since, $V = 0$ and the polarity of the back emf remains unchanged, the armature current during dynamic braking is determined from the equation

$$I_a = -\frac{E_b}{R_a + R_{db}} = -\frac{K\phi\omega}{R_a + R_{db}} \tag{2.12}$$

The negative (–) sign shows that the armature current is reversed. The dynamic braking torque can be expressed as

$$T_{db} = K\phi I_a = -\frac{(K\phi)^2\,\omega}{R_a + R_{db}} \tag{2.13}$$

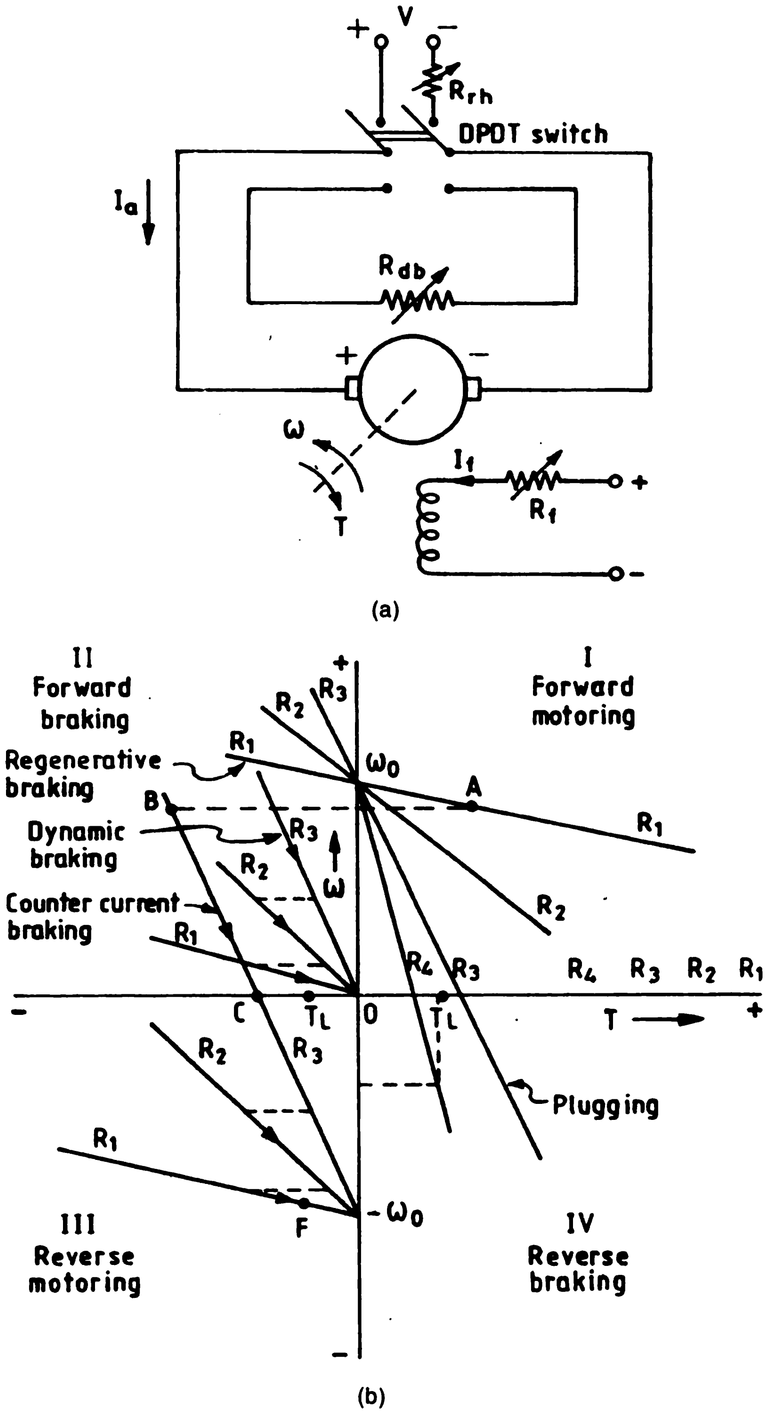


Fig. 2.6 Circuit and speed-torque characteristics of dc shunt motors: (a) Basic circuit for dynamic braking; (b) Speed-torque characteristics depicting four-quadrant operation.

or

$$\omega = -\frac{(R_a + R_{db})T_{db}}{C^2} \tag{2.14}$$

It may be seen from Eq. (2.14) that the speed-torque curve under dynamic braking conditions is a straight line, lying in the second quadrant and finishing at the origin of the coordinate axes. A family of dynamic braking characteristics for different values of the external resistance is shown in Fig. 2.6b. The characteristics show that the hardness decreases with increase in the value of the armature circuit resistance. During dynamic braking, the excitation of the field winding is maintained by separate excitation. Braking resistance is decreased in steps as in starting, otherwise braking torque would be very small at low speeds. The maximum and minimum limits of currents during braking may be taken to be around 200% and 100% of rated current respectively. The resistance can be calculated graphically from the speed-torque characteristics. This method of braking is applied in non-reversible drives.

2.4.2 Counter Current Braking (Plugging)

If the polarity of the supply voltage to the armature is reversed while the motor is running, the type of braking in the motor is termed *counter current braking* or *plugging* (Fig. 2.7).

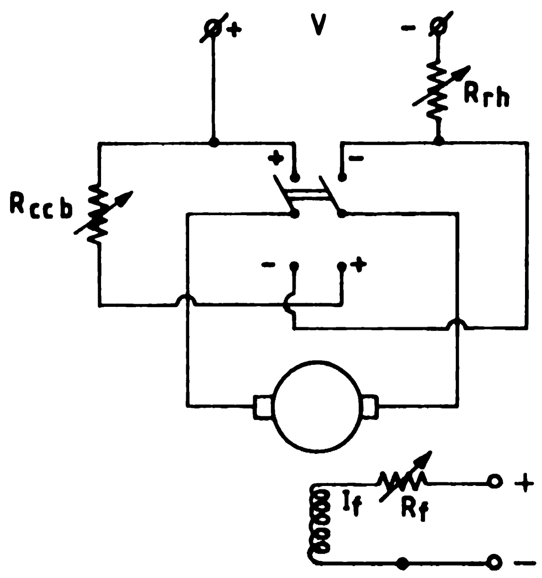


Fig. 2.7 Basic circuit for counter current braking (plugging) of a dc shunt motor.

Motor emf has the same polarity as the reversed input voltage. The armature current is given by

$$I_a = \frac{V + E_b}{R} \tag{2.15}$$

where R = total resistance in the armature circuit. As $E_b \approx V$, in order to limit the counter current braking current to rated value, $R = 2V/I_L$.

With the reversal of the supply voltage, the normal direction of rotation is also reversed, shifting the point corresponding to no load speed from $+\omega_0$ to $-\omega_0$. The direction of the motor torque is also reversed. The speed-torque characteristic equation may be written as

$$\omega = -\frac{V}{C} + \frac{TR}{C^2} \tag{2.16}$$

In Fig. 2.6b, the motor is operating at a speed corresponding to the point A in quadrant I. The armature voltage is reversed with a suitable external resistance in the armature circuit. The operating point is transferred from the point A to the point B in quadrant II on the characteristic curve drawn according to Eq. (2.16). As braking torque is developed, the motor speed drops down along the characteristic curve BC in quadrant II. If the purpose of braking is to stop the motor, the supply has to be switched off as soon as the speed is zero at the point C. If the supply is not switched off, the motor accelerates in the opposite direction in quadrant III. Since the direction of rotation has changed, the motor emf also reverses, thus opposing the supply voltage. The drive now produces the motoring torque and becomes stable at the point F on the natural characteristic corresponding to the load torque, when the plugging resistance is decreased.

A situation identical to the condition of counter current braking arises during lowering of an overhauling load. Suppose that a hoist motor is started to raise a load (Fig. 2.8a) and as such the motor is operating on the characteristic AB at the point A in the quadrant I with load torque T_L at the shaft (Fig. 2.8c).

During motoring, the current is given by

$$I_a = \frac{V - E_b}{R}$$

Now, if load is increased and when load torque T_L becomes equal to short circuit torque T_{sc}, the motor stalls. The speed and back emf drop to zero and the armature current is determined from

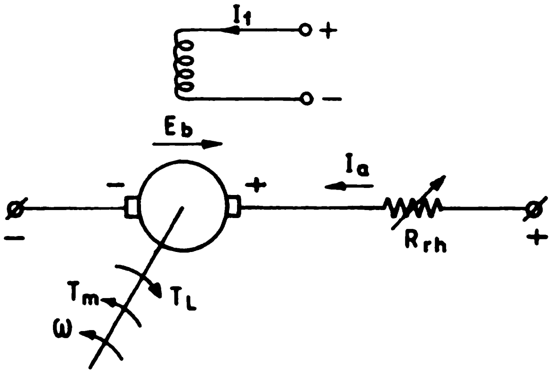
$$I_a = \frac{V}{R}$$

Up to this point, the maximum motoring torque occurs.

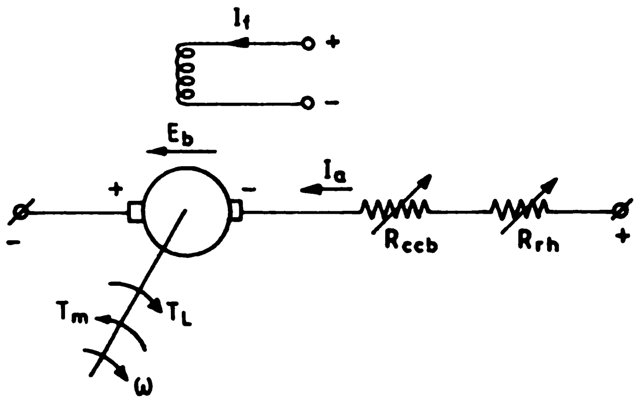
On further increase of the load torque, the motor experiences an overhauling torque. The armature rotates in the opposite direction due to gravity action (Fig. 2.8b). At a torque, T_{L2}, the load descends with a speed corresponding to the point C lying in the fourth quadrant (Fig. 2.8c). The armature rotates in the direction opposite to the normal, corresponding to the applied voltage. The polarity of the field remains unchanged. The motor emf reverses so that

$$I_a = \frac{V + E_b}{R}$$

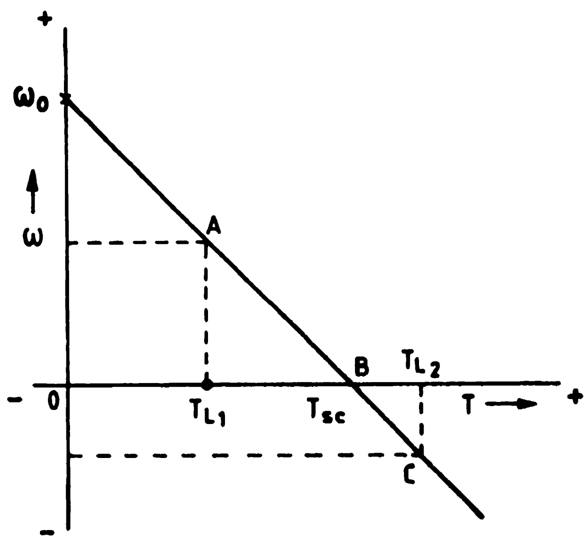
Greater torque and more rapid stop are obtained by plugging. Current is drawn from the supply. This energy together with the stored energy has to be dissipated in a resistance. A relay must be used to open the circuit, when the speed drops to zero.



(a) Circuit for normal hoist operation.



(b) Circuit for counter current braking (plugging) during lowering of an overhauling load.



(c) Speed-torque characteristic showing motoring and plugging operations.

Fig. 2.8 Circuits and speed-torque characteristics of dc shunt motors.

2.4.3 Regenerative Braking

When a motor runs at a speed higher than the ideal no load speed corresponding to the field current or flux, the back emf exceeds the applied voltage, V. The machine then operates as a generator in parallel to the supply. The energy is fed back into the supply. This type of situation also exists, when the field current is increased and the motor speed does not decrease instantly, i.e. the speed exceeds the no load speed for the new value of the field current or flux. The back emf now exceeds the supply voltage and thus, the machine acts as a generator as long as this condition is fulfilled. This is called regeneration and the form of braking is called the *regenerative braking*.

$$\text{The armature current} = \frac{V - E_b}{R} = -\frac{E_b - V}{R} = -I_a$$

The motor torque consequently changes its sign and the motor develops a braking torque.

$$T = -K\phi I_a = -T_b, \text{ at } \omega > \omega_0$$

The speed-torque characteristic equation takes the form

$$\omega = \frac{V}{K\phi} + \frac{RT_b}{(K\phi)^2} \tag{2.17}$$

The speed-torque characteristic at $T_b = 0$, passes through the point corresponding to ideal no load speed with any external resistance in the armature circuit. The braking characteristics have the same gradient as that of the motoring characteristics. Thus, the regenerative braking characteristics are just a continuation of the motoring characteristics from quadrant I to quadrant II (Fig. 2.6b).

Example 2.3

A 500 V, 45 kW, 600 rpm dc shunt motor has full load efficiency of 90%. The field resistance is 200 Ω and the armature resistance is 0.2 Ω. The field current is maintained constant. Armature reaction and brush drop may be neglected. Calculate the rated armature current and hence, find the speed under each of the following conditions at which the machine develops an electromagnetic torque equal to the rated value.

(a) Regenerative braking: no external resistance.

(b) Plugging: external resistance of 5.5 Ω inserted.

(c) Dynamic braking: external resistance of 2.6 Ω inserted.

Solution

$P_o = 45$ kW, $\eta = 0.9$, $V = 500$ V, $N = 600$ rpm, $R_f = 200\ \Omega$, $R_a = 0.2\ \Omega$

$$\text{Motor input current at full load } I_i = \frac{P_o}{\eta \cdot V} = \frac{45 \times 10^3}{0.9 \times 500} = 100 \text{ A}$$

$$I_f = \frac{V}{R_f} = \frac{500}{200} = 2.5 \text{ A}$$

Armature current (full load) $I_L = I_i - I_f = 100 - 2.5 = 97.5$ A

$$\omega = \frac{2\pi N}{60} = \frac{2\pi \times 600}{60} = 62.83 \text{ rad/s}$$

$$E_b = V - I_L R_a = 500 - (97.5 \times 0.2) = 480.5 \text{ V}$$

emf coefficient = torque coefficient = $C = \frac{E_b}{\omega} = \frac{480.5}{62.83} = 7.65$ V · s/rad or N · m/A

Motor electromagnetic torque at rated current, $T_e = CI_L = 7.65 \times 97.5 = 745.9$ N · m

Nominal shaft torque of the motor $T_o = \frac{P_o}{\omega} = \frac{45 \times 10^3}{62.83} = 716.2$ N · m

No-load torque $T' = T_e - T_o = 745.9 - 716.2 = 29.7$ N · m

Electromagnetic torque equal to the rated value,

$$T_i = T_e + T' = 745.9 + 29.7 = 775.6 \text{ N} \cdot \text{m}$$

(a) *Regenerative braking*

$$N = \frac{60}{2\pi}\omega = \frac{60}{2\pi}\left(\frac{V}{C} + \frac{T_b R_a}{C^2}\right) = \frac{60}{2\pi}\left(\frac{500}{7.65} + \frac{775.6 \times 0.2}{(7.65)^2}\right) = 650 \text{ rpm}$$

(b) *Plugging*

Plugging current $I_{pl} = \frac{T_{pl}}{C} = \frac{775.6}{7.65} = 101.4$ A

Total resistance in the armature circuit during plugging,

$$R = R_a + R_{pl} = 0.2 + 5.5 = 5.7\ \Omega$$

$$N = \frac{60}{2\pi}\omega = \frac{60}{2\pi}\left(\frac{I_{pl}R - V}{C}\right) = \frac{60}{2\pi}\left(\frac{101.5 \times 5.7) - 500}{7.65}\right) = 97.3 \text{ rpm}$$

(c) *Dynamic braking*

$$R = R_a + R_{db} = 0.2 + 2.6 = 2.8\ \Omega$$

$$N = \frac{60}{2\pi}\omega = \frac{60}{2\pi}\left(\frac{T_{db}R}{C^2}\right) = \frac{60}{2\pi}\left(\frac{775.6 \times 2.8}{(7.65)^2}\right) = 354.4 \text{ rpm}$$

2.4.4 Determination of Starting, Dynamic Braking and Plugging Resistance for DC Shunt Motors

Calculation of starting resistance

Appearing in Fig. 2.9 is a typical circuit diagram of a starter for dc shunt motors. The problem

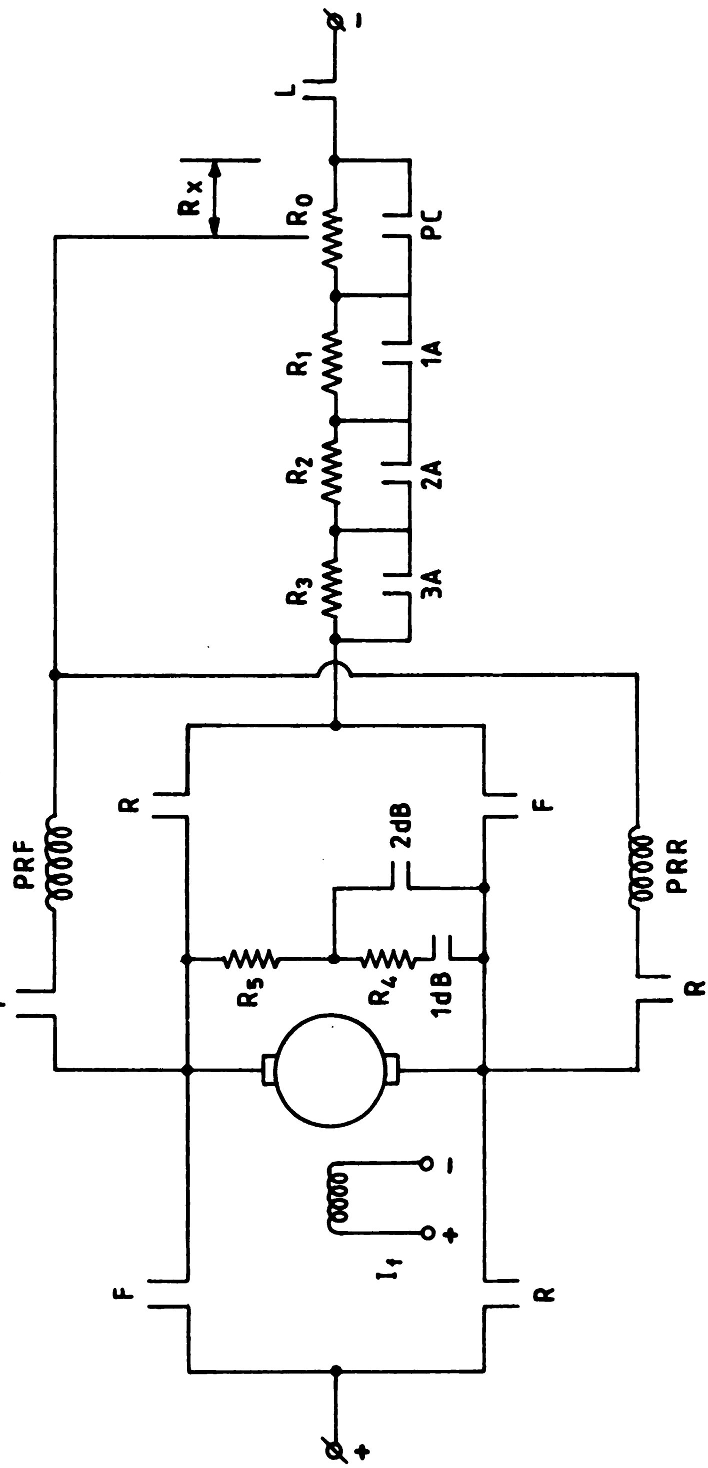


Fig. 2.9 Circuit diagram of a starter for dc shunt motors.

concerning determination of values of starting resistances can be solved graphically (Fig. 2.10). The maximum and minimum currents (torques) during starting are 180% and 100% for a three-step starting. With the total resistance ($R_1 + R_2 + R_3$) included at starting ($\omega = 0$), the characteristic starts at the point G corresponding to 180% of nominal torque ($I_a = 1.8\ I_{nom}$) and ends at the point P corresponding to the ideal no load ($I_a = 0$). The speed rises corresponding to the point E, when the current drops to I_{nom} ($T = T_{nom}$). At this instant, the accelerating contactor 1A closes, cutting off the resistance R_1. The lower value of accelerating torque is found to be 100% of nominal torque. The characteristic changes from E to H (corresponding to 180% of nominal torque) instantaneously with total resistance equal to ($R_2 + R_3$) and then proceeds along the line HP. The process as stated above is repeated and the various characteristics JP and QP can be drawn as shown in Fig. 2.10. The motor is finally transferred to the natural characteristic. The construction is successful, if the horizontal line CQ touches the natural characteristic at Q. The FA represents $R_{nom} = V_{nom}/I_{nom}$. The AB represents the drop across the armature resistance R_a at nominal (full load) current. Similarly, BC, CD and DE represent the drops across resistances, R_3, R_2 and R_1, respectively at the same current. Each resistance is found as a percentage of R_{nom}.

$$FA = R_{nom}; \quad R_1 = \frac{ED}{FA} R_{nom}; \quad R_2 = \frac{DC}{FA} R_{nom}; \quad R_3 = \frac{CB}{FA} R_{nom}; \quad R_a = \frac{AB}{FA} R_{nom}$$

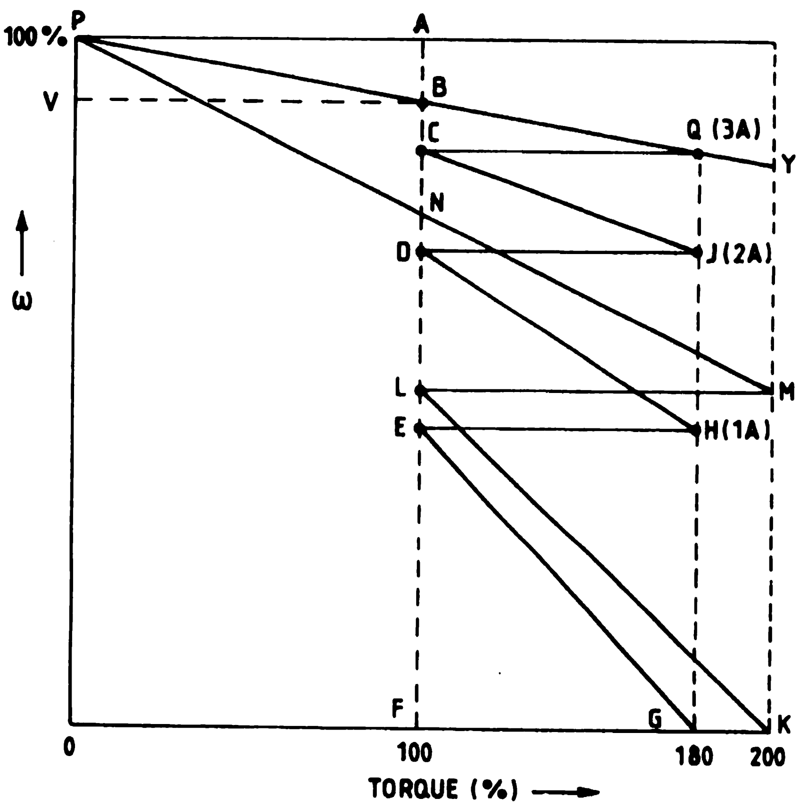


Fig. 2.10 DC shunt motor—calculation of starting resistance.

Calculation of dynamic braking resistance

Suppose that a two-step dynamic braking is to be accomplished. The characteristics *QP* and *MP* corresponding to $(R_4 + R_5)$ and R_5, respectively are drawn in Fig. 2.10. The peak dynamic braking torque is assumed to be 200%.

$$R_4 = \frac{LN}{FA} R_{nom}; \qquad R_5 = \frac{NB}{FA} R_{nom}$$

Calculation of plugging resistance

$$R_0 + R_1 + R_2 + R_3 + R_a = \frac{V + E_{max}}{I_{pl}}$$

where

E_{max} = maximum possible no load voltage, and

I_{pl} = permitted plugging current

For shunt motor, $V \approx E_{max}$

If I_{pl} = 200% of full load current, then

$$R_0 + R_1 + R_2 + R_3 + R_a = \frac{2V}{2I_{nom}} = R_{nom}$$

or

$$R_0 = R_{nom} - (R_1 + R_2 + R_3 + R_a)$$

$$R_x = \left(\frac{V}{V + E_{max}}\right) R_{nom} = \frac{1}{2} R_{nom} = 50\% \, R_{nom}$$

To ensure reliable pick up, when the motor is started, it is set for a pick-up voltage 20% lower than the voltage across the relay, when the motor is at rest. For a shunt motor,

$$V_{pick\text{-}up} = 0.4 \times V$$

Example 2.4

A dc motor (shunt with series stabilizing winding) has the following specifications:

Voltage = 220 V, Output = 14 kW, Speed = 710 rpm,

$I_{nom} = 75$ A, $R_a = 0.26\ \Omega$, $R_{stab} = 0.0037\ \Omega$

Find the resistance to be included for starting, dynamic braking, and plugging. The upper and lower limits of torque are 180% and 100%, respectively.

Solution

$$R_{nom} = \frac{V_{nom}}{I_{nom}} = \frac{220}{75} = 2.933\ \Omega$$

$$R_a = \frac{0.26}{2.933} \times 100 = 8.86\%\ R_{nom}$$

(a) *Starting resistance*
Upper limit of torque = 180% rated torque
Lower limit of torque = 100% rated torque
With reference to Fig. 2.10, we have

$$FA = R_{nom} = 2.933\ \Omega$$

$$R_1 = \frac{ED}{FA} R_{nom} = 0.25 \times 2.933 = 0.733\ \Omega$$

$$R_2 = \frac{DC}{FA} R_{nom} = 0.14 \times 2.933 = 0.411\ \Omega$$

$$R_3 = \frac{CB}{FA} R_{nom} = 0.075 \times 2.933 = 0.22\ \Omega$$

(b) *Dynamic braking resistance*

$$R_4 = \frac{LN}{FA} R_{nom} = 0.25 \times 2.933 = 0.733\ \Omega$$

$$R_5 = \frac{NB}{FA} R_{nom} = 0.162 \times 2.933 = 0.475\ \Omega$$

(c) *Plugging resistance*
FE represents the plugging resistance, R_0

$$R_0 = R_{nom} - (R_1 + R_2 + R_3 + R_a) = 2.933 - (0.733 + 0.411 + 0.22 + 0.26) = 1.31\ \Omega$$

$$V_{pick\text{-}up} = 0.4 \times V = 0.4 \times 220 = 88\ \text{V}$$

2.5 SPEED-TORQUE CHARACTERISTICS OF DC SERIES MOTORS

The speed equation of a series motor is written as

$$\omega = \frac{V - I_a R}{K\phi} \tag{2.18}$$

where R = total resistance in the armature circuit.

The motor torque may be expressed as $T = K\phi\, I_a$. In a series motor, the flux ϕ depends upon armature current I_a as shown in Fig. 2.11.

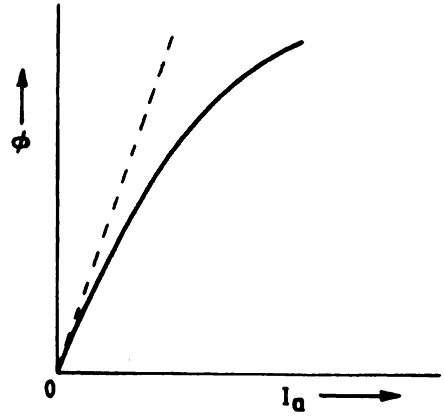


Fig. 2.11 Typical magnetization curve of dc series motors.

If, for simplification, $\phi - I_a$ relationship is assumed to be linear, then, $\phi = K_1 I_a$. Now

$$T = K\phi I_a = KK_1(I_a)^2$$

or

$$I_a = \sqrt{\frac{T}{KK_1}}$$

Using the above equations in Eq. (2.18), we obtain

$$\omega = \frac{V}{K\phi} - \frac{I_a R}{K\phi} = \frac{V}{KK_1 I_a} - \frac{R}{KK_1} = \frac{V}{\sqrt{KK_1}} \cdot \frac{1}{\sqrt{T}} - \frac{R}{KK_1} = \frac{A}{\sqrt{T}} - B \tag{2.19}$$

If the flux is assumed to be constant due to saturation of the magnetic circuit, then

$$\omega = \frac{V}{K\phi} - \frac{I_a R}{K\phi} = C - DT \, (\because \quad T \propto I_a) \tag{2.20}$$

where C and D are constants. It follows that the speed-torque characteristic takes the form as shown in Fig. 2.12, having the ordinate as its asymptote. A sharp rise in speed is due to weakening of the field at light loads.

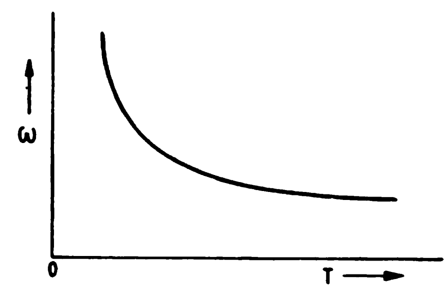


Fig. 2.12 Typical speed-torque characteristic of dc series motors.

2.5.1 Construction of Speed-Torque Characteristics

It would be wrong to assume that a series motor works in the unsaturated region of its magnetization curve. Rather, series motors should not be used for drives, where the shaft torque may fall below 15–20% of rated value. So, for calculation of the speed-torque characteristic, saturation of magnetic circuit of the machine should be considered. However, the flux cannot be considered to be constant.

It is customary to resort to the grapho-analytical method in which calculations are based on natural characteristics, $N = f_1(I)$ and $T = f_2(I)$, supplied by the manufacturer or obtained experimentally. The characteristic curves are given on the basis of per unit values so that they may be applied in general. The typical speed and torque curves on per unit values are given in Fig. 2.13. Speed and torque curves in absolute values are obtained from the universal characteristics and nameplate data of the motor.

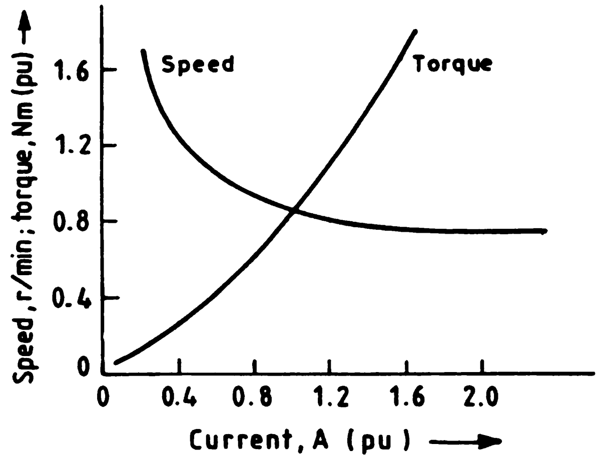


Fig. 2.13 Universal speed and torque curves on per unit (pu) values basis for dc series motors.

2.5.2 Determination of Starting Resistance for DC Series Motors

The starting resistance can be determined graphically from the natural characteristic of the motor without using the rheostatic characteristics. The following procedure is followed (Fig. 2.14).

1. Draw the natural characteristic of the motor.

2. Select the maximum and minimum limits of the starting current, i.e. I_I and I_{II} at which switching will take place from one step to another.

3. Determine R_I and R_{II} from $\frac{V}{I_I}$ and $\frac{V}{I_{II}}$, respectively.

4. Draw the vertical dotted line Ah to the left of the origin at a distance OA proportional to the motor internal resistance R_m, using a suitable scale.

5. Draw horizontal projections from the points of I_I and I_{II} on the natural characteristic curve to the vertical dotted line Ah intersecting it at points h and g.

6. Cut off the distances, $Oa = R_I = \frac{V}{I_I}$ and $Ok = R_{II} = \frac{V}{I_{II}}$ to the same resistance scale.

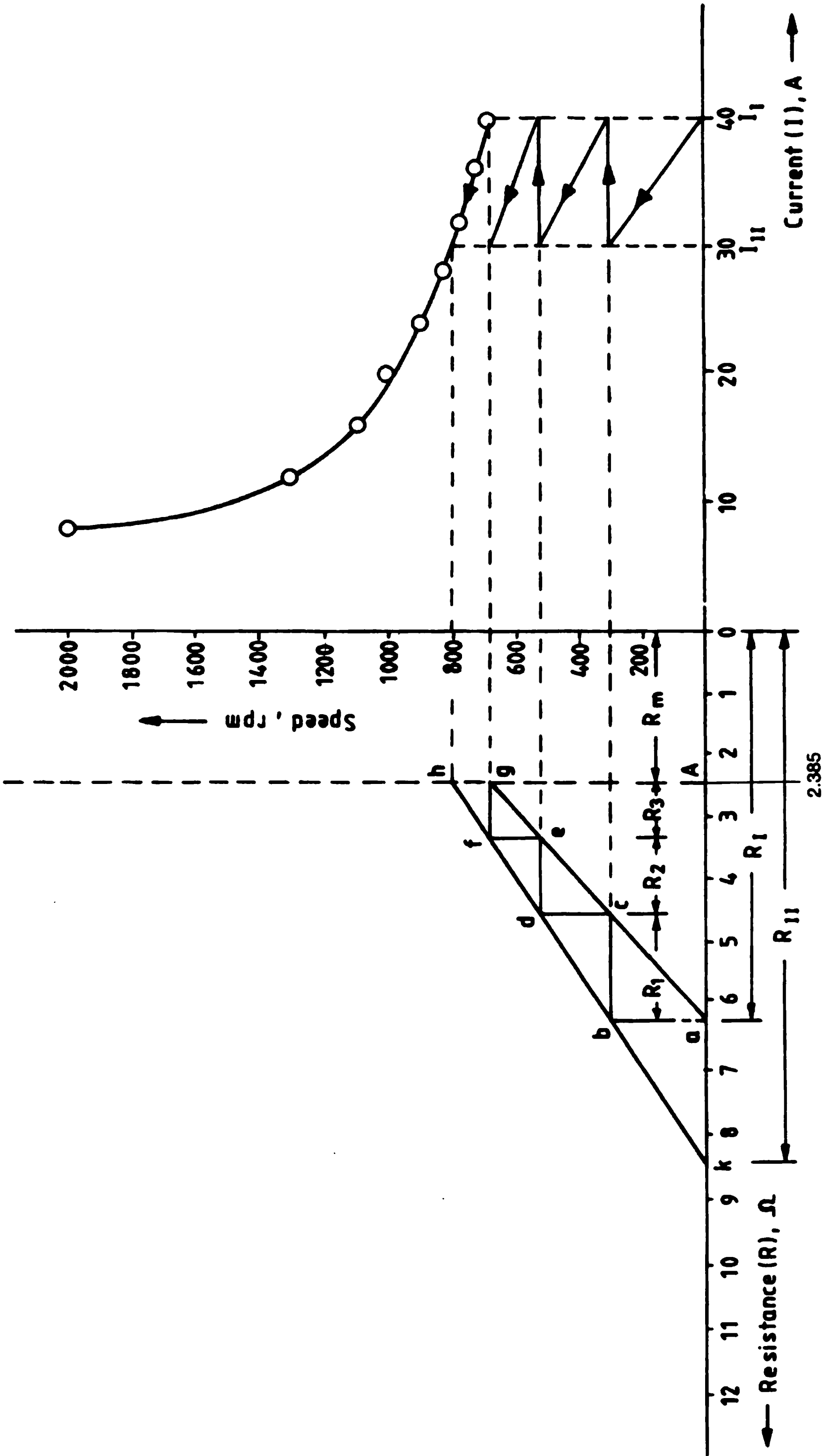


Fig. 2.14 Graphical determination of starting resistance of dc series motors.

7. Join *kh* and *ag*, which give linear relationship between the speed of the motor and the resistance in the armature circuit of the motor at constant armature current. This follows from the equation

$$\omega = \frac{V - I_a R}{K\phi} = A - BR \tag{2.21}$$

8. Draw the vertical line *ab*, which intersects the line *kh* at *b*. From this point draw a horizontal line, which intersects the line *ag* at *c*. The lines *cd, de, ef* and *fg* are drawn in the same way. The construction is successful, if the last horizontal line passes through the point *g*.

9. If the construction is not successful, select different values of I_I and/or I_{II} and repeat the same procedure until the construction is successful.

10. The resistances, R_1, R_2 and R_3 are obtained from the lengths *bc*, *de* and *fg*, respectively.

Example 2.5

Determine the three-step resistance for starting a dc series motor having the following ratings:

$P_{nom} = 3.73$ kW, $V_{nom} = 250$ V, $I_{nom} = 20$ A, $N_{nom} = 1000$ rpm

Solution

The natural speed-torque characteristic is first plotted. The data for plotting this curve are obtained by making use of Fig. 2.13.

$i_{nom} = 1.0$, $I_{nom} = 20$ A, $v_{nom} = 1.0$, $N_{nom} = 1000$ rpm

The currents and speeds for different per unit values are calculated from

$$I = iI_{nom} \text{ and } N = v\,N_{nom}$$

The results are given in Table 2.1, from which the natural characteristic is plotted as in Fig. 2.14.

Table 2.1

i, pu	0.4	0.6	0.8	1.0	1.2	1.4	1.6	1.8	2.0
I, A	8	12	16	20	24	28	32	36	40
v, pu	2.0	1.3	1.09	1.0	0.9	0.82	0.78	0.72	0.68
N, rpm	2000	1300	1090	1000	900	820	780	720	680

The maximum starting current I_I is taken to be $2I_{nom} = 40$ A. After a few trials, it is found that the lower limit of the current I_{II} is equal to $1.5I_{nom} = 30$ A. Then, we have

$$R_I = \frac{V}{I_I} = \frac{250}{40} = 6.25\ \Omega$$

$$R_{II} = \frac{V}{I_{II}} = \frac{250}{30} = 8.33\ \Omega$$

The graph in Fig. 2.14 is drawn according to the procedure outlined above. The resistances R_1, R_2 and R_3 are obtained as

$$R_1 = 1.5\ \Omega, \quad R_2 = 1.2\ \Omega, \quad R_3 = 0.9\ \Omega$$

2.5.3 Types of Braking (Applications in DC Cranes)

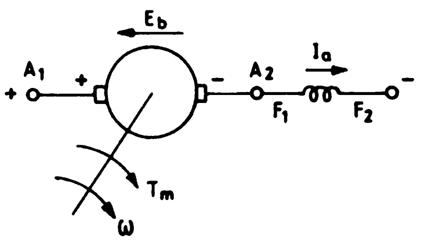
The braking of a dc series motor may be accomplished in two ways: (i) dynamic or rheostatic braking, and (ii) counter current braking or plugging. Regenerative braking is not possible with a series motor, as the back emf of the motor cannot exceed the supply voltage.

In dynamic braking with self-excitation, the supply to the motor is switched off and then the armature circuit including the series field winding is connected across a resistor (Figs. 2.15a and b). Care should be taken such that the excitation is not reversed during the changeover.

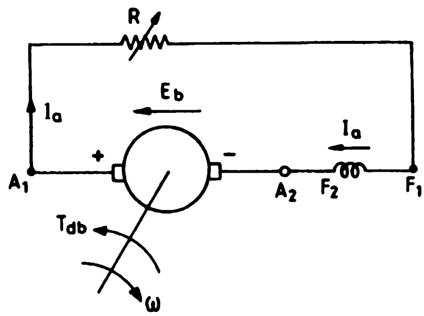
The dynamic braking torque is

$$T_{db} = -\frac{(K\phi)^2\,\omega}{R_a + R_{se} + R_{db}}$$

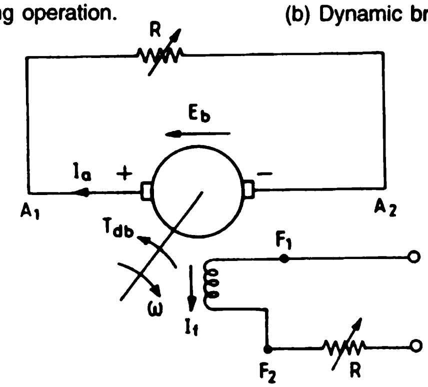
Unlike the separately excited or shunt motor, the flux ϕ is dependent on the armature current I_a. When braking is initiated, the current is high, thus, resulting in increased value of flux, and the torque is also high, being approximately proportional to square of the current. The speed-torque characteristics for dynamic braking are in the second quadrant (Fig. 2.16). At this instant, the driven unit may experience objectionable shocks due to a large value of braking torque. The machine now runs as a self-excited generator. The scheme with separate excitation is devoid of this drawback, as the series field current can be controlled (Fig. 2.15c). The principle is the same as in the case of separately excited dc motors described in Section 2.3.1.



(a) Motoring operation.



(b) Dynamic braking with self-excitation.



(c) Dynamic braking with separate excitation.

Fig. 2.15 Basic circuits used for dc series motors.

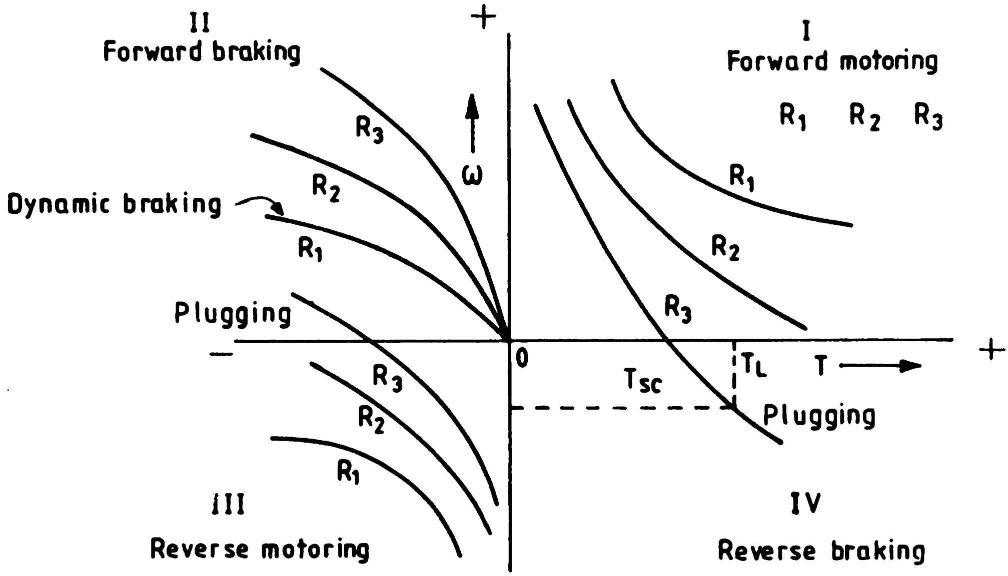


Fig. 2.16 Speed-torque characteristics of dc series motors under different operating conditions.

Dynamic braking is applied in cranes in case of power failure, while the crane is in operation. On occurrence of power failure, the armature circuit is connected in series with the field circuit and a suitable external resistance (Fig. 2.17). The motor is overhauled by the load on the hook due to gravity. It acts as a self-excited generator. The stored energy is dissipated as heat in the armature circuit resistance and dynamic braking action takes place.

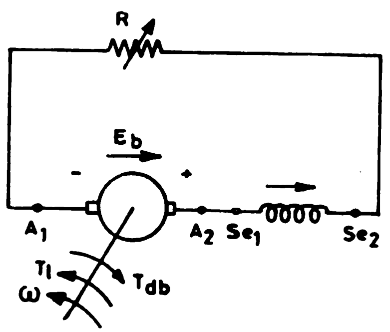


Fig. 2.17 Circuit for dynamic braking of a dc series motor for lowering of a loaded cage in a crane.

The voltage equation is

$$-E_b = I_a(R_a + R_{se} + R_{db})$$

As the speed is in the reverse direction, the direction of the generated or back emf, E_b also reverses. So, the series field need not be reversed in this case as the current continues to flow in the same direction. The speed-torque characteristics lie in the fourth quadrant (Fig. 2.16). This may also be termed as emergency braking. As stated earlier, braking torque cannot be properly controlled. Other methods are available to control braking torque, but are not described here.

The counter current braking (plugging) is accomplished by reversing the polarity of the armature terminals. When the armature current is reversed, the direction of the field current is maintained in the same direction by changing the series field connection as in the case of dynamic braking shown in Fig. 2.15b. The speed-torque curves for counter current braking are shown in Fig. 2.18. A suitable value of external resistance is inserted into the armature circuit during counter current braking to limit the armature current.

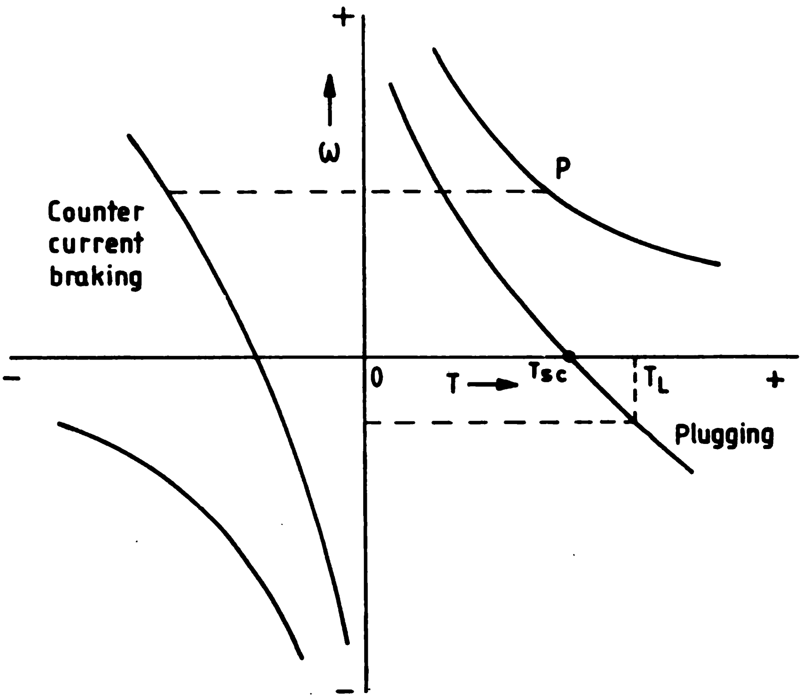


Fig. 2.18 Speed-torque characteristics illustrating counter current braking (plugging) of dc series motors.

As in the case of dynamic or rheostatic braking, this type of braking (plugging) can be applied, when the loaded cage is being lowered. If the motoring characteristic with external resistance connected in series is extended into the fourth quadrant as shown in Fig. 2.18, it can be observed that reverse braking takes place with braking torque (positive) developed in the machine, but the speed is negative, i.e. in the reverse direction. The load torque T_L exceeds the torque T_{sc} developed by the machine with the external resistance in series, at standstill or rest. The series field connection is the same as shown in Fig. 2.17 for the case of dynamic braking during lowering. It may also be noted that the direction of armature current is positive, i.e. same as the torque (positive) developed with the field current or flux being positive, but the induced (back) emf is negative, as the speed is negative. The supply voltage is in the positive direction.

Example 2.6

A dc series motor of ratings given in Example 2.5, is to lower a load with counter current braking at an armature current of $I_a = 1.5\, I_{nom}$ and speed of 200 rpm in one case and 300 rpm in another. Determine the additional resistance needed to be inserted in the armature circuit.

Solution

From the natural characteristic of the motor drawn in Example 2.5 (Fig. 2.14), at

$$I_1 = 1.5I_{nom} = 30 \text{ A}, \qquad \text{the speed, } N_1 = 800 \text{ rpm}$$

$$\omega_1 = \frac{2\pi N_1}{60} = \frac{2\pi \times 800}{60} = 83.8 \text{ rad/s}$$

Then, at current $I_1 = 30$ A, we have

$$K\phi = \frac{V - I_a R_m}{\omega_1} = \frac{250 - 30 \times 2.385}{83.8} = 2.13 \text{ V} \cdot \text{s/rad}$$

The emf during counter current braking,

$$E_b = K\phi\omega_{br}$$

where ω_{br} is the angular speed of the armature during counter current braking.

The armature current during counter current braking is given by

$$I_a = \frac{V + E_b}{R_m + R_{rh}}$$

or

$$R_{rh} = \frac{V + E_b}{I_a} - R_m$$

At the lowering speed of $N_2 = 200$ rpm, we have

$$\omega_2 = \frac{2\pi \times 200}{60} = 20.95 \text{ rad/s}$$

$$R_{rh2} = \frac{250 + 2.13 \times 20.95}{30} - 2.385 = 7.435 \ \Omega$$

At the lowering speed of $N_3 = 300$ rpm, we get

$$\omega_3 = \frac{2\pi \times 300}{60} = 31.42 \text{ rad/s}$$

$$R_{rh3} = \frac{250 + 2.13 \times 31.42}{30} - 2.385 = 8.18 \ \Omega$$

2.6 SPEED-TORQUE CHARACTERISTICS OF DC COMPOUND MOTORS

2.6.1 Cumulative Compound Motor

In this motor, the mmf of the series field winding assists that of the shunt field winding. That is why the characteristics lie in intermediate positions with respect to dc separately excited or shunt motors, and series motors. By adjusting the strength (number of turns) of the windings, any desired characteristic can be obtained. The torque equation is

$$T = K(\phi_{sh} + \phi_{se})I_a$$

At no load, the flux ϕ_{sh} is produced only by the shunt field winding. With the increase in load torque, the flux, ϕ_{se} produced by the series field winding, being a function of I_a (same as series field current), is added to ϕ_{sh}. So, a cumulatively compound motor always produces a higher torque than that produced by a separately excited or shunt motor for the same armature current, as shown in Fig. 2.19a. The non-linearity of the compound motor characteristics is due to the saturation of the magnetic circuit, as the series field mmf cannot produce significant amount of flux with the increase

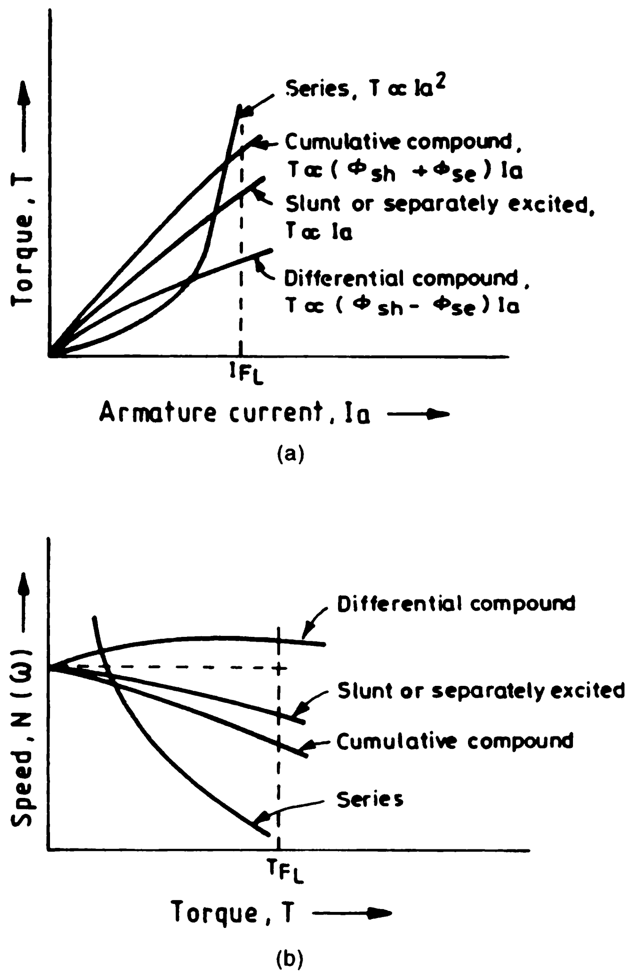


Fig. 2.19 Characteristics of different types of dc motors: (a) Torque-current characteristics; (b) Speed-torque characteristics.

in load torque. The speed-torque characteristics of a cumulatively compound motor can be expressed as

$$\omega = \frac{V}{K\phi_{sh}} - \frac{T(R_a + R_{se})}{K^2(\phi_{sh} + \phi_{se})^2}$$

Thus, the speed of a cumulative compound motor decreases at a faster rate than that of a separately excited or shunt motor with the increase in load torque, as shown in Fig. 2.19b.

Unlike series motors, the compound motors have ideal no load speed given by

$$\omega_0 = \frac{V}{K\phi_{sh}}$$

However, since the field flux $(\phi_{sh} + \phi_{se})$ changes with load and saturation of the core occurs, it is very difficult to use the equations to compute the speed-torque characteristics. Like series motors, universal speed-current and torque-current characteristics (supplied by the manufacturer), as shown in Fig. 2.20, are used to compute the speed-torque characteristics.

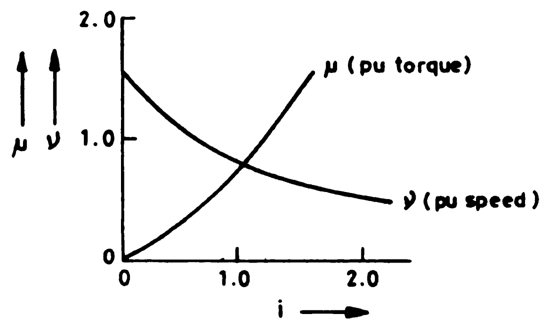


Fig. 2.20 Speed and torque curves as a function of current for dc compound (cumulative) motors.

Cumulative compound motors are widely used for individual drives like lifts, winches, etc. where series motor characteristics are required with safe no load speed. These motors are particularly suited to heavy intermittent loads such as presses.

2.6.2 Types of Braking

Electric braking of compound motors may take place in three ways:

(a) Regenerative braking.

(b) Dynamic or rheostatic braking, and

(c) Counter current braking, or plugging.

The characteristic curves under braking conditions are shown in Fig. 2.21.

In regenerative braking, the direction of the current in the armature and series field winding reverses. This causes demagnetizing effect on the machine. When the speed reaches the ideal no load speed, the series field winding is shunted to prevent demagnetization. So, the regenerative braking curves are straight lines lying in the second quadrant.

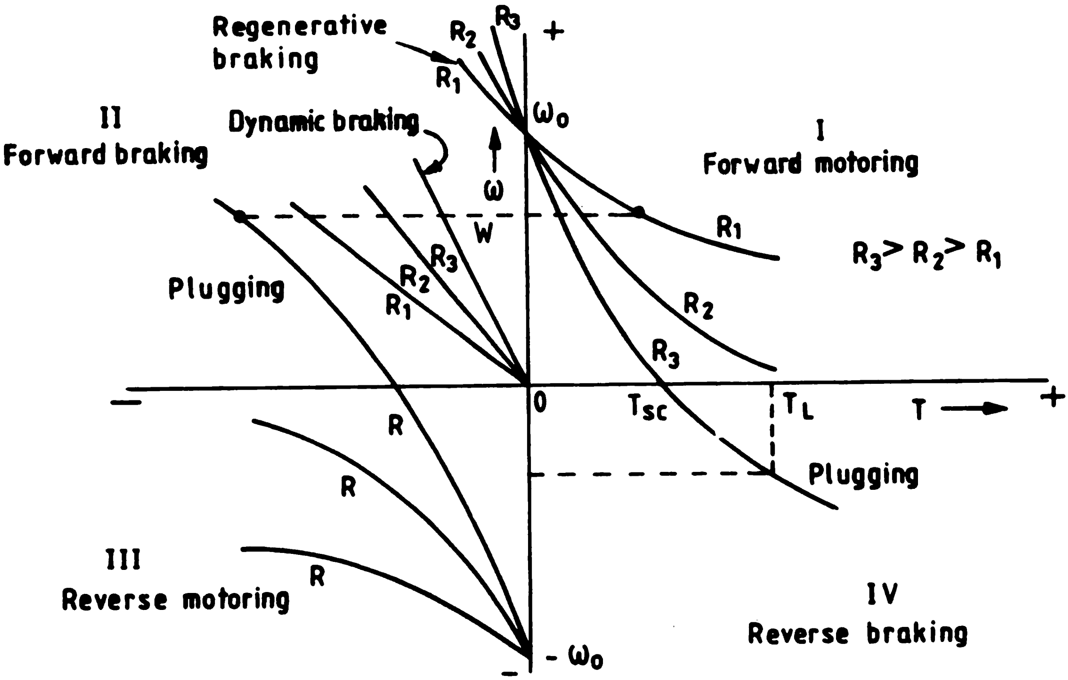


Fig. 2.21 Motoring and braking characteristics of cumulative compound motors.

The dynamic braking curves are also straight lines, because only under the dynamic braking condition, the shunt field winding is excited.

The counter current braking (plugging) occurs, when the load overpowers the power developed in the armature. These curves are continuation of motoring curves from the first quadrant into the fourth quadrant. They are not straight lines.

2.6.3 Differential Compound Motor

The mmf of the series winding opposes that of the shunt winding. The differential compound motor is rarely used because of the disadvantage that the field is weakened too much while taking excessive armature current. This may cause instability of motor operation. At starting, the current is so excessive that the mmf of the series winding may be exceed that of the shunt winding causing the armature to rotate in the opposite direction. If a level compound generator is connected for motor operation, differential action takes place.

2.7 TORQUE-SPEED CHARACTERISTICS OF INDUCTION MOTORS

2.7.1 Torque Equation, Typical Starting Performance

The three-phase induction motor is represented by an approximate equivalent circuit as shown in Fig. 2.22.

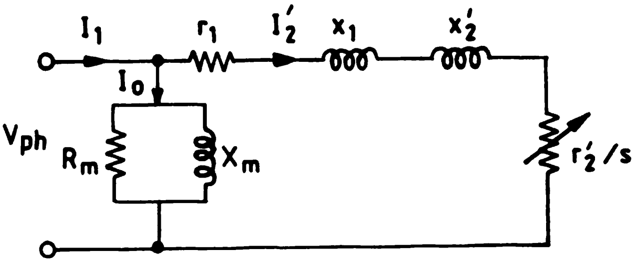


Fig. 2.22 Approximate equivalent circuit of the three-phase induction motor.

Supply frequency (Hz) $= f$

Synchronous angular speed $\omega_s = 2\pi f \left(\frac{2}{p}\right)$ rad/s

where p = no. of poles.

Rotor angular speed (rad/s) $= \omega_r$

The slip is defined as

$$s = \frac{\omega_s - \omega_r}{\omega_s} \tag{2.22}$$

The rotor current referred to the stator is,

$$I_2' = \frac{V_{ph}}{\sqrt{\left(r_1 + \frac{r_2'}{s}\right)^2 + (x_1 + x_2')^2}} \tag{2.23}$$

where

V_{ph} = voltage per phase applied to the stator,

r_1 and r_2' are the stator and rotor resistance per phase respectively referred to the stator, and

x_1 and x_2' are the stator and rotor leakage reactance per phase respectively referred to the stator.

The developed torque in N · m is

$$T = \frac{3(I_2')^2 \frac{r_2'}{s}}{\omega_s} \tag{2.24}$$

$$= \frac{3 V_{ph}^2 r_2'}{s\omega_s \left[\left(r_1 + \frac{r_2'}{s}\right)^2 + (x_1 + x_2')^2\right]} \tag{2.25}$$

The gross mechanical power output (W), $P_{mech} = \omega_r T$

The total power input to the rotor, $P_{ag} = \omega_s T$

The rotor copper loss, $P_2 = (\omega_s - \omega_r)T = s\omega_s T = sP_{ag}$

The power output, $P_0 = P_{ag} - P_2 = (1 - s)P_{ag}$

Maximum torque

The maximum torque occurs, when $\frac{dT}{ds} = 0$.

The corresponding slip at maximum torque is

$$s_m = \frac{\pm r_2'}{\sqrt{(r_1)^2 + (x_1 + x_2')^2}} \tag{2.26}$$

The maximum torque is

$$T_m = \frac{3V_{ph}^2}{2\omega_s\left[r_1 \pm \sqrt{(r_1)^2 + (x_1 + x_2')^2}\right]} \tag{2.27}$$

The positive sign (+) in Eqs. (2.26) and (2.27) corresponds to the motoring operation. If $1 \geq s > 0$, the induction machine acts as motor. If $s > 1$, plugging operation takes place.

The negative sign (−) in above equations corresponds to the generating operation. The induction motor runs at negative slip, when the rotor speed is greater than the synchronous speed. Regenerative braking takes place under this condition.

Dividing Eq. (2.25) by Eq. (2.27), we get

$$\frac{T}{T_m} = \frac{2r_2'\left[r_1 \pm \sqrt{(r_1)^2 + (x_1 + x_2')^2}\right]}{s\left[\left(r_1 + \frac{r_2'}{s}\right)^2 + (x_1 + x_2)^2\right]}$$

Substituting r_2' by $\pm s_m \cdot \sqrt{(r_1)^2 + (x_1 + x_2')^2}$ and putting $\alpha = \frac{r_1}{r_2'}$, the above equation may be simplified to the form

$$T = \frac{2T_m(1 + \alpha s_m)}{\left(\frac{s}{s_m} + \frac{s_m}{s} + 2\alpha s_m\right)} \tag{2.28}$$

If the stator resistance and leakage reactance are neglected, Eq. (2.28) can be simplified as

$$T = \frac{2T_m}{\left(\frac{s}{s_m} + \frac{s_m}{s}\right)} \tag{2.29}$$

where $s_m = \frac{r_2'}{x_2'}$.

Figure 2.23 shows a typical torque-speed characteristic of the three-phase induction motor operating at constant voltage and frequency.

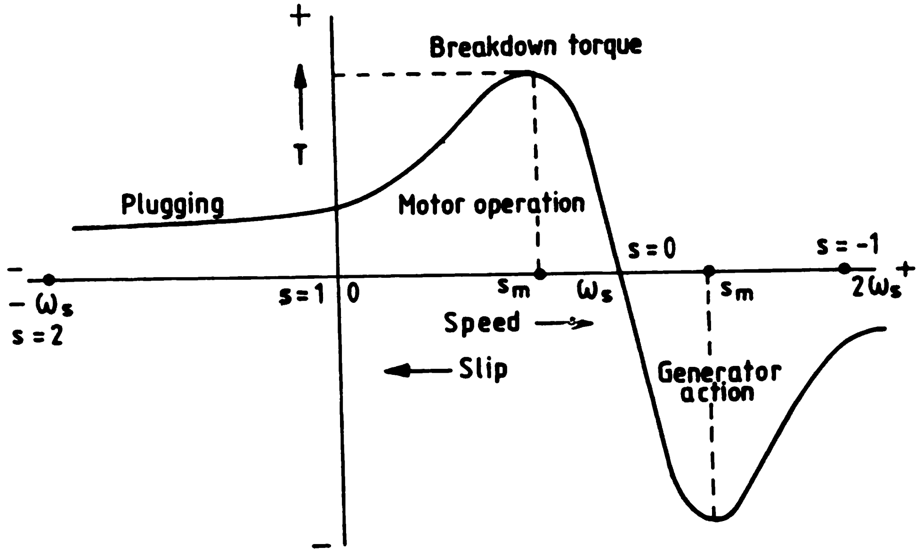


Fig. 2.23 Typical torque-speed characteristic of three-phase induction motors.

It may be noted that the values of s_m are identical for both motoring and generating operations. It is also clear from Eq. (2.27) that during generating operation, the maximum torque is greater than that during motoring.

Power-slip characteristics

The total internal mechanical power developed is

$$P_{\text{mech}} = (1-s)P_{\text{ag}} = 3(I_2')^2(1-s)\frac{r_2'}{s} = \frac{3V_{\text{ph}}^2 r_2'(1-s)}{s\left[\left(r_1 + \frac{r_2'}{s}\right)^2 + (x_1 + x_2')^2\right]}$$

$$= \frac{3V_{\text{ph}}^2 r_2' \cdot s(1-s)}{[(r_1 s + r_2')^2 + s^2(x_1 + x_2')^2]} \tag{2.30}$$

The effect of no-load current I_0 on the primary impedance drop has been neglected so far. But for an induction motor, the no-load current I_0 is large, often as much as 35–40% of the full-load current. Therefore, I_0 should not be neglected for derivation of the equations for torque and internal mechanical power. The effect of no-load current on the stator impedance drop may be taken into account by replacing V_{ph} by

$$V_e = V_{\text{ph}} - I_0 x_1$$

in the equations for torque and power, while the subtraction can be made arithmetically. It is sufficiently accurate in most cases.

Example 2.7

The maximum torque of a three-phase induction motor is twice the full-load torque and the starting torque is equal to the full-load torque. Calculate the full-load slip. Neglect the stator impedance.

Solution

From Eq. (2.29), we can write

$$T_{fl} = \frac{2T_m}{\left(\dfrac{s_{fl}}{s_m} + \dfrac{s_m}{s_{fl}}\right)}$$

or

$$\frac{s_{fl}}{s_m} + \frac{s_m}{s_{fl}} = \frac{2T_m}{T_{fl}} = 4, \text{ as } T_m = 2T_{fl}$$

Also

$$T_{st} = \frac{2T_m}{\left(\dfrac{1}{s_m} + s_m\right)}, \quad \text{as } s = 1 \text{ at starting } (\because\ \omega_r = 0)$$

From the given condition

$$T_{fl} = T_{st}$$

Then

$$\frac{1}{s_m} + s_m = \frac{2T_m}{T_{st}} = 4$$

or

$$s_m = 0.268 = 26.8\%$$

Now

$$\frac{s_{fl}}{s_m} + \frac{s_m}{s_{fl}} = 4$$

or

$$\frac{s_{fl}}{0.268} + \frac{0.268}{s_{fl}} = 4$$

Therefore

$$s_{fl} = 0.0718 = 7.18\%$$

2.7.2 Starters for Slip-ring Induction Motors

The slip-ring induction motor gives high starting torque with low starting current. These motors are well suited for high inertia loads which take a long time to accelerate. Smooth acceleration and speed control are obtained using the rotor resistance starters. High starting torque with low starting current demands high rotor resistance. On the other hand, the rotor resistance has to be low for low full-load slip, high efficiency and moderate rotor heating. Therefore, the basic control system would need insertion of an extra resistance into the rotor circuit at the time of starting. This resistance is progressively reduced as the motor speeds up and ultimately cut out (Fig. 2.24). Some resistance is retained in the circuit, if speed control is required.

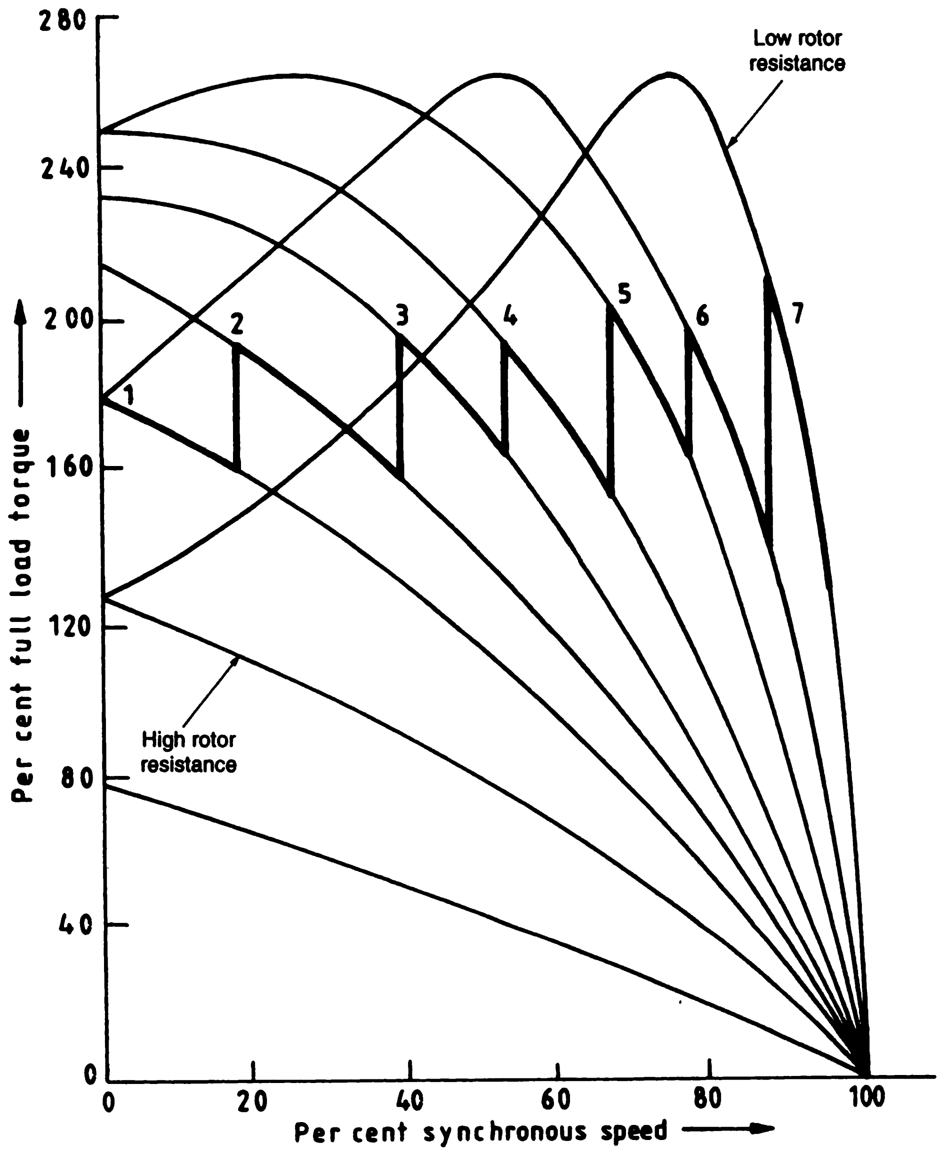


Fig. 2.24 Torque-speed characteristics of a slip-ring induction motor showing the effect of variation of rotor resistance.

The rotor resistances of a slip-ring induction motor are basically of two types—liquid and metallic. The liquid starter allows smooth starting. Metallic resistances are used for all ratings. The air-cooled starting resistance is used for small and medium size motors. The oil-immersed resistance is used for higher ratings. It is sometimes more economical to use a separate rotor resistance for starting and speed regulation. The cost of this type of motor is high. A comparison of starting performances with various types of starters of both single- and three-phase induction motors is given below in Table 2.2.

Table 2.2 Comparison of Starting Performances

$$\text{Current ratio } (\%) = \frac{\text{Starting current}}{\text{Full-load current}} \qquad \text{Torque ratio } (\%) = \frac{\text{Starting torque}}{\text{Full-load torque}}$$

Type of induction motor (IM)	*Method of starting*	*Starting performances*	
		Current ratio (%)	*Torque ratio (%)*
Single-phase IM			
Split phase	DOL	150	600
Capacitor start	DOL	300	350
Capacitor start/Capacitor run	DOL	100	400
Three-phase IM			
Cage rotor	DOL	700	150
	Star/Delta	250	50
	Auto-transformer 80% tap	400	100
Inverted-T cage rotor (Sash bar)	DOL	500	125
	Star/Delta	150	40
	Auto-transformer 80% tap	300	80
Double-cage rotor	DOL	650	250
	Star/Delta	210	80
	Auto-transformer 80% tap	400	150
Three-cage rotor	DOL	525	250
	Star/Delta	160	80
	Auto-transformer 80% tap	300	150
Wound rotor (slip-ring)	Rotor resistance	200	200

Devices for facilitating starting

Centrifugal clutches or coupling devices of hydraulic, magnetic or electric type, may be fitted between the motor and the load to reduce the load torque during starting. As a result either the starting time is reduced or the starting current or torque is lowered, facilitating the use of low cost squirrel-cage motors instead of the more costly types of motors.

Soft-start

A three-phase ac regulator using back-to-back connection of thyristors or triacs in the supply lines for phase-controlled operation as described in Section 4.5 can be used for soft-start of a three-phase induction motor.

At start, 180° delay is provided. Phasing forward is controlled at a predetermined rate or by monitoring acceleration so that the starting process is without mechanical and electrical shock. The current limit controller may be incorporated to maintain the starting current at a preset value. However, in any case, the frequency remains constant.

2.7.3 Types of Braking

As in case of a dc motor, the electrical braking of an induction motor can be accomplished in three ways:

(a) Plugging,

(b) Dynamic or rheostatic braking, and

(c) Regenerative braking.

Plugging (counter current braking)

In this method, the braking torque is produced by interchanging any two supply terminals, so that the direction of rotation of the rotating magnetic field is reversed with respect to the rotation of the motor. The electromagnetic torque developed provides the braking action and brings the rotor to a quick stop.

Let s be the slip for the motoring operation with respect to the forward rotating field. Then the corresponding slip after reversal of the rotating magnetic field will be $(2 - s)$.

$$s \text{ (during motoring)} = \frac{\omega_s - \omega_r}{\omega_s}$$

or

$$\omega_r = (1 - s)\omega_s$$

$$\text{The slip during plugging } s_{pl} = \frac{-\omega_s - \omega_r}{-\omega_s} = \frac{\omega_s + \omega_r}{\omega_s} = 2 - s$$

If the rotor is of slip-ring (wound rotor) type, external resistance may be inserted into the rotor circuit to limit the plugging current. The motor should be disconnected from the supply at the time when the speed drops to zero. Otherwise, the motor would continue to run in the opposite direction in reverse motoring mode.

The counter current braking condition can also be set up, when load torque T_{L1} exceeds the stalling torque T_m (Fig. 2.25). The rotor rotates in the direction opposite to that of the rotating magnetic field. External resistance is inserted into the rotor circuit of the slip-ring induction motor in order to limit the plugging current and develop the required plugging torque.

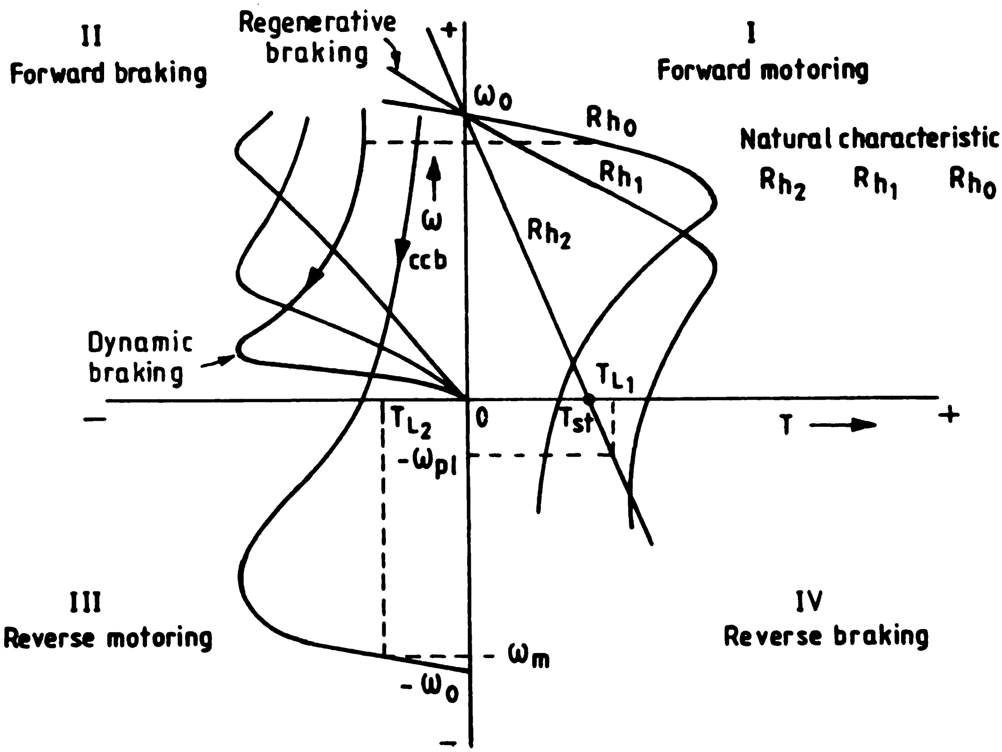


Fig. 2.25 Typical speed-torque characteristics of induction motors under different operating conditions.

Example 2.8

A 400 V, three-phase, 50 Hz, four-pole cage induction motor has the following equivalent circuit parameters:

$$r_1 = 0.1\ \Omega, \quad x_1 = 0.4\ \Omega, \quad r_2' = 0.1\ \Omega, \quad x_2' = 0.4\ \Omega, \quad x_m = 14.0\ \Omega$$

The motor was operating on full-load at slip = 0.05, when the two stator terminals were suddenly interchanged. Calculate the primary current and the braking torque immediately after application of plugging.

Solution

$$r_1 = r_2' = 0.1\ \Omega, \quad x_1 = x_2' = 0.4\ \Omega, \quad x_m = 14.0\ \Omega, \quad f = 50\text{ Hz}, \quad p = 4$$

$$V_{ph} = \frac{400}{\sqrt{3}} = 230\text{ V}, \qquad N_s = \frac{120f}{p} = \frac{120 \times 50}{4} = 1500\text{ rpm}$$

$$\omega_s = \frac{2\pi}{60} N_s = \frac{2\pi \times 1500}{60} = 157.1\text{ rad/s}$$

Magnetizing current $I_m = \dfrac{V_{ph}}{jx_m} = \dfrac{230 + j0.0}{0.0 + j14} = -j16.43 = 16.43 \angle -90° \text{ A}$

$s = 0.05, \quad 2 - s = 2 - 0.05 = 1.95, \quad \dfrac{r_2'}{2-s} = \dfrac{0.1}{1.95} = 0.0513\ \Omega$

$$z = \left(r_1 + \frac{r_2'}{2-s}\right) + j(x_1 + x_2') = (0.1 + 0.0513) + j(0.4 + 0.4) = 0.1513 + j0.8$$

$$= 0.8142 \angle 79.3°\ \Omega$$

Rotor current referred to the stator $I_2' = \dfrac{V_{ph}}{z} = \dfrac{230 \angle 0°}{0.8142 \angle 79.3°} = 282.5 \angle -79.3° \text{ A}$

Primary current $= I_1 = I_m + I_2' = 16.43 \angle -90° + 282.5 \angle -79.3° = 299.0 \angle -79.88 \text{ A}$

Braking torque $T_b = \dfrac{3(I_2')^2}{\omega_s} \cdot \dfrac{r_2'}{2-s} = \dfrac{3 \times (282.5)^2 \times 0.0513}{157.1} = 78.19 \text{ N} \cdot \text{m}$

Regenerative braking

An induction motor is subjected to regenerative braking, if the rotor speed exceeds the synchronous speed of the motor. Under regenerative braking, the machine acts as an induction generator returning energy to the supply and taking only the reactive power for excitation. When the rotor speed exceeds the synchronous speed, the slip becomes negative. The regenerative braking characteristic is the continuation of the motoring characteristic into the upper part of quadrants II/IV (Fig. 2.25). The maximum regenerative braking torque is higher than the maximum motoring torque.

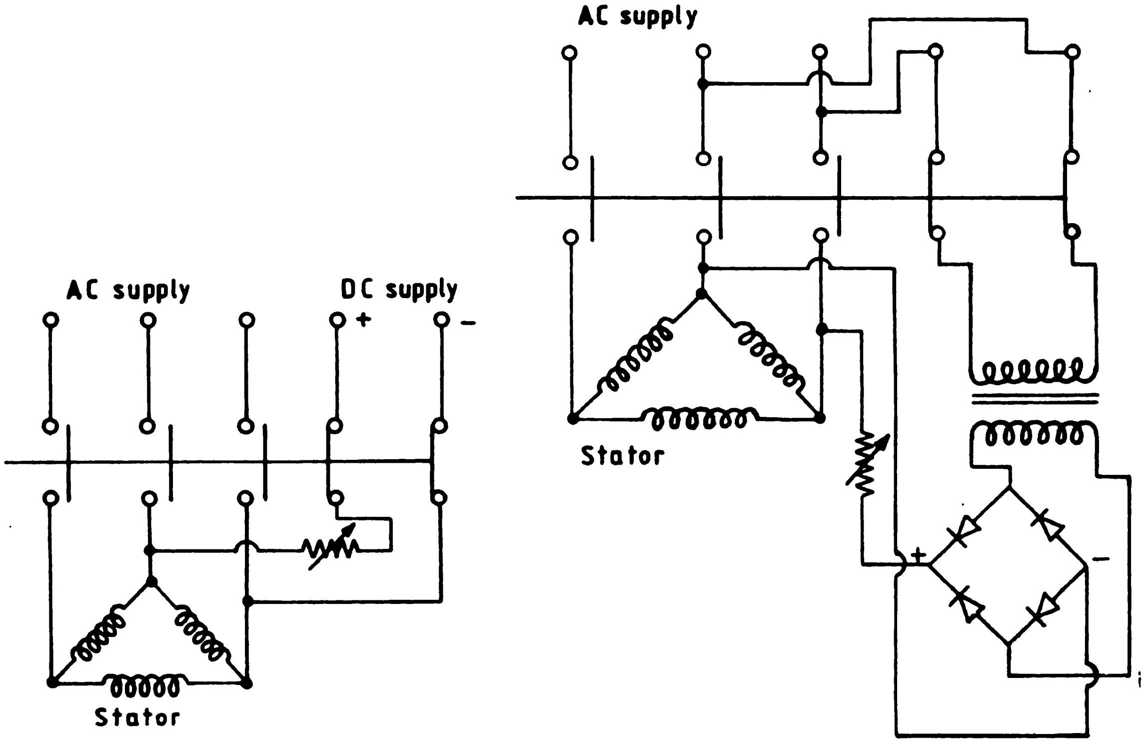
When the number of poles of a pole-changing induction motor is changed in the ratio 1 : 2, for example, four-pole to eight-pole, regenerative braking takes place immediately after the changeover, till the lower steady state speed is reached.

Dynamic braking

Dynamic braking of the induction motor is achieved by disconnecting the stator windings from the ac supply and connecting it to the dc supply (Fig. 2.26). When the machine is motoring, the stator magnetic field rotates in the same direction as the rotor. During dynamic braking, the stator magnetic field is stationary, as it is fed from dc supply, and the rotor continues to rotate in this stationary field. Alternating current is thus induced in the rotor windings. This current produces a rotating magnetic field, which rotates at the same speed as the rotor, but in the direction opposite to that of the rotor such that it becomes stationary with respect to the stator. The rotor current, therefore, flows in the direction opposite to that corresponding to the motoring action. Hence, a braking torque is produced.

Initially, the frequency of the rotor current corresponds to nearly the synchronous speed. The relative speed between the stationary magnetic field and the rotating rotor at any speed is given by

$$(1 - s)\omega_s = S\omega_s$$



(a) DC source is available. (b) DC voltage is obtained using a rectifier.

Fig. 2.26 Circuit diagram for dc dynamic braking of induction motors.

Therefore, as the speed drops, the frequency will also reduce and become zero at standstill. When the motor accelerates, the rotor frequency decreases and comes to zero at synchronous speed. The conditions in the rotor during dynamic braking are very similar to those when the motor accelerates from standstill.

The magnitude of the braking torque depends upon dc excitation, the rotor speed and the rotor resistance. Figure 2.25 illustrates dynamic braking torque vs. speed characteristics, which lie in the lower side of the quadrant II.

Calculation of dynamic braking torque

The induction motor operates in generating mode during dynamic braking. The dynamic braking torque may be determined in the following way, which is based on cylindrical rotor theory.

The stator windings are fed from a balanced three-phase supply, while the rotor is rotated at synchronous speed. The voltage is varied and plotted against the excitation current, I_{ac} (Fig. 2.27). This gives the magnetization characteristic of the motor.

Now, the excitation voltage at the stator terminals at any speed ω_r, when the motor is operating in generating mode during dynamic braking (Fig. 2.28), may be deduced as

$$E_s = \frac{\omega_r}{\omega_s} E_0 = (1 - s)E_0 \tag{2.31}$$

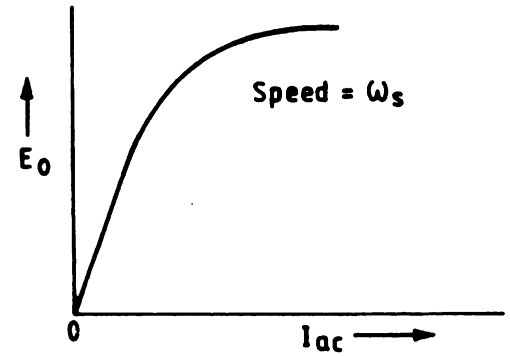


Fig. 2.27 Typical magnetization characteristic of induction motors.

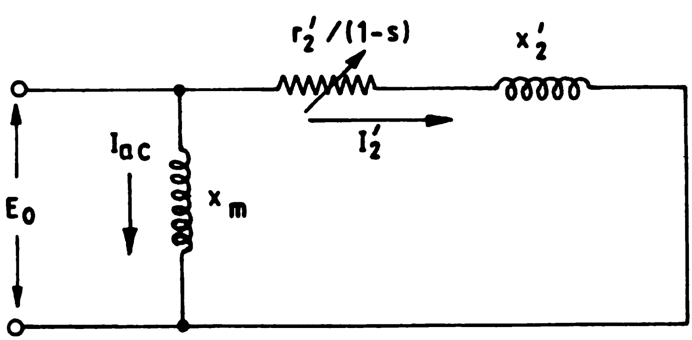


Fig. 2.28 Equivalent circuit of the induction motor during dynamic braking.

where

E_s = induced (excitation) voltage in the stator at any speed ω_r,

E_0 = induced (excitation) voltage in the stator at synchronous speed ω_s, and

Slip during motoring $s = \dfrac{\omega_s - \omega_r}{\omega_s} = 1 - \dfrac{\omega_r}{\omega_s}$

Synchronous reactance = $j(1 - s)x_s$

Slip during dynamic braking $S = 1 - s$

I_2' = stator induced current, which flows in the direction opposite to that of the current in motoring mode.

Then, electromagnetic torque (braking)

$$T_b = -\frac{3(1-s)E_0 I_2' \cos\phi_2}{\omega_r} = -\frac{3E_0^2\left(\dfrac{r_2'}{1-s}\right)}{\omega_s\left[\left(\dfrac{r_2'}{1-s}\right)^2 + (x_s)^2\right]} \tag{2.32}$$

since

$$I_2' = \frac{(1-s)E_0}{\sqrt{(r_2')^2 + (1-s)^2 (x_s)^2}} \tag{2.33}$$

and

$$\cos \phi_2 = \frac{r_2'}{\sqrt{(r_2')^2 + (1-s)^2 (x_s)^2}} \tag{2.34}$$

But

$$E_0 = I_{ac} x_m \text{ and } x_s = x_m + x_2'$$

Hence

$$T_b = -\frac{3(I_{ac} x_m)^2 \left(\frac{r_2'}{1-s}\right)}{\omega_s \left[\left(\frac{r_2'}{1-s}\right)^2 + (x_s)^2\right]} = -\frac{3(I_{ac} x_m)^2 \left(\frac{r_2'}{S}\right)}{\omega_s \left[\left(\frac{r_2'}{S}\right)^2 + (x_s)^2\right]} \tag{2.35}$$

Differentiating with respect to S and equating to zero, the maximum braking torque is obtained as

$$(T_b)_{\max} = -\frac{3(I_{ac} x_m)^2}{\omega_s \cdot 2(x_m + x_2')} \tag{2.36}$$

The critical dynamic braking slip at which maximum dynamic torque occurs is

$$S_m = \frac{r_2'}{(x_m + x_2')}$$

After simple transformation, the torque equation (Eq. 2.36) takes the form

$$T_b = \frac{2(T_b)_{\max}}{\left(\frac{S}{S_m} + \frac{S_m}{S}\right)} \tag{2.37}$$

In order to determine the electromagnetic torque during dynamic braking, it is necessary to find the equivalent ac excitation from dc excitation. This equivalence is obtained by equating the mmfs created by equivalent alternating current and the direct current. The excitation depends upon the type of the stator connection. The four types of connections widely used are shown in Fig. 2.29.

$$\text{The amplitude of mmf produced by } I_{ac} = \frac{3}{2}\sqrt{2}\, I_{ac} K_w N_1 = \frac{3}{2}\sqrt{2}\, I_{ac} N_1'$$

For connection (a) in Fig. 2.29, Phase A and Phase B carry equal currents, but in opposite directions. So, the corresponding space mmf vectors are displaced by 60°.

$$\text{The amplitude of the resultant mmf} = \sqrt{3}\, I_{dc} N_1'$$

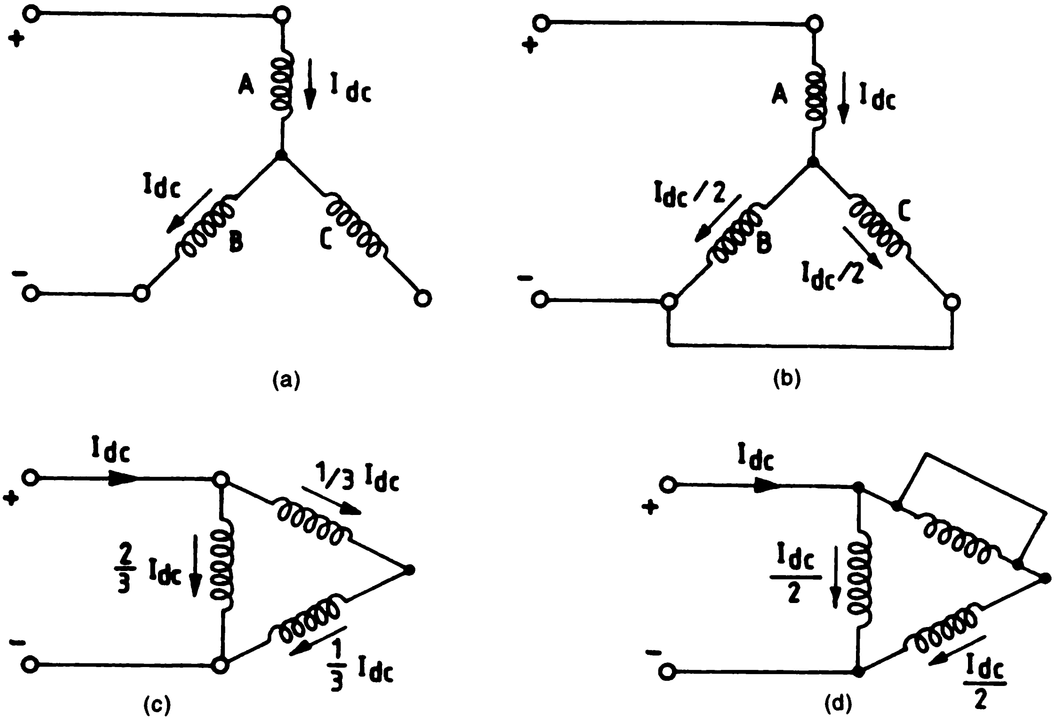


Fig. 2.29 Different types of stator winding connections in induction motors used for dc dynamic braking.

Equating amplitudes due to I_{ac} and I_{dc}, we get

$$\frac{3}{2}\sqrt{2}\, I_{ac}\, N_1' = \sqrt{3}\, I_{dc} N_1'$$

or

$$I_{ac} = \sqrt{\frac{2}{3}}\, I_{dc}$$

Similarly, for connection (b) in Fig. 2.29

$$I_{ac} = \frac{1}{\sqrt{2}}\, I_{dc}$$

For connection (c)

$$I_{ac} = \frac{\sqrt{2}}{3}\, I_{dc}$$

and for connection (d)

$$I_{ac} = \frac{1}{\sqrt{6}}\, I_{dc}$$

Example 2.9

A 500 V, three-phase, 50 Hz, eight-pole, star-connected induction motor has the following parameters of its equivalent circuit.

$r_1 = 0.13\ \Omega$, $r_2 = 0.32\ \Omega$, $x_1 = 0.6\ \Omega$, $x_2 = 1.48\ \Omega$, $r_m = 250\ \Omega$, $x_m = 20\ \Omega$

The full-load slip is 5%. The effective stator to rotor turns ratio per phase is 1/1.57. The machine is to be braked from full-load speed by changing the stator connections and inserting an external rotor circuit resistance, which in primary terms is 1.5 Ω per phase (referred to the stator). Determine the initial braking torque, when the stator is disconnected from the ac supply and dc is fed into two of its terminals. Determine also the required dc excitation and the braking torque for counter current braking. Neglect mechanical losses and use the approximate equivalent circuit.

Solution

$$r_1 = 0.13\ \Omega,\quad r_2 = 0.32\ \Omega,\quad x_1 = 0.6\ \Omega,\quad x_2 = 1.48\ \Omega,$$

$$a = \frac{1}{1.57},\quad V_{ph} = \frac{500}{\sqrt{3}} = 288.67\ \text{V},\quad f = 50\ \text{Hz},\quad p = 8$$

$$N_s = \frac{120f}{p} = \frac{120 \times 50}{8} = 750\ \text{rpm},\quad \omega_s = \frac{2\pi}{60} N_s = \frac{2\pi \times 750}{60} = 78.54\ \text{rad/s}$$

Rotor resistance referred to the stator $r_2' = a^2 r_2 = \dfrac{0.32}{(1.57)^2} = 0.13\ \Omega$

Rotor leakage reactance referred to the stator $x_2' = a^2 x_2 = \dfrac{1.48}{(1.57)^2} = 0.6\ \Omega$

The equivalent circuit of the motor at slip $s = 0.05$ and the equivalent circuit during dynamic braking from full-load speed are shown in Figs. 2.30(a) and (b), respectively.

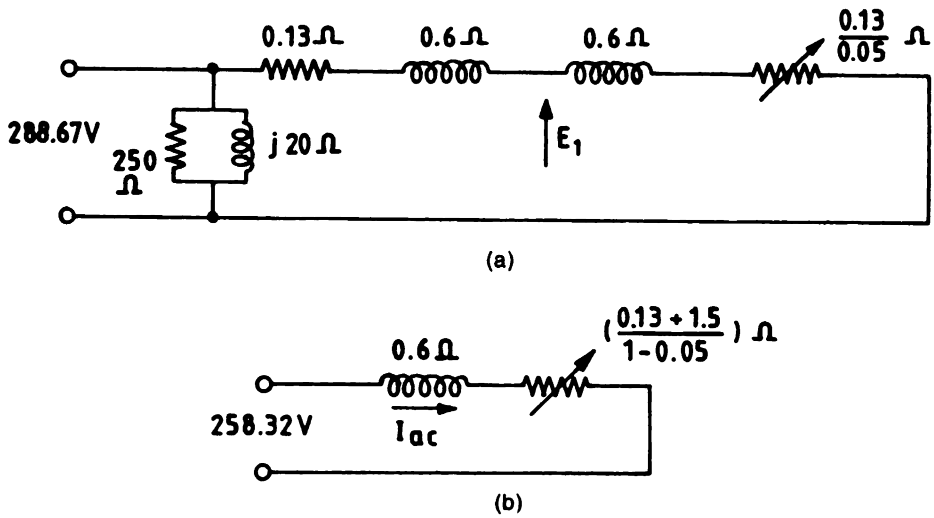


Fig. 2.30 Equivalent circuits (Example 2.9): (a) Motoring; (b) Dynamic braking.

$$\frac{r_2'}{s} = \frac{0.13}{0.05} = 2.6\ \Omega, \qquad x_1 + x_2' = 0.6 + 0.6 = 1.2\ \Omega$$

Now

$$E_1 = V_{ph} \frac{\sqrt{\left(\frac{r_2'}{s}\right)^2 + (x_2')^2}}{\sqrt{\left(r_1 + \frac{r_2'}{s}\right)^2 + (x_1 + x_2')^2}} = \frac{288.67 \times \sqrt{(2.6)^2 + (0.6)^2}}{\sqrt{(0.13 + 2.6)^2 + (1.2)^2}} = 258.3\ \text{V}$$

Dynamic braking slip $S = 1 - s = 1.0 - 0.05 = 0.95$

$$\frac{r_2' + R'}{S} = \frac{0.13 + 1.5}{0.95} = \frac{1.63}{0.95} = 1.716\ \Omega$$

$$I_{ac} = \frac{E_1}{\sqrt{\left(\frac{r_2' + R'}{S}\right)^2 + (x_2')^2}} = \frac{258.3}{\sqrt{\left[(1.716)^2 + (0.6)^2\right]}} = 142.1\ \text{A}$$

Initial dynamic braking torque

$$T_{db} = \left(\frac{3(I_{ac})^2}{\omega_s}\right)\left(\frac{r_2' + R'}{S}\right) = \frac{3 \times (142.1)^2 \times 1.716}{78.54} = 1323.54\ \text{N} \cdot \text{m}$$

Required dc excitation $I_{dc} = \sqrt{\frac{3}{2}}\, I_{ac} = 1.2247 \times 142.1 = 174\ \text{A}$

During counter current braking (plugging), initial slip $s_{pl} = 2 - s = 2 - 0.05 = 1.95$

$$\frac{r_2' + R'}{2 - s} = \frac{1.63}{1.95} = 0.836\ \Omega$$

Therefore, the initial current referred to the stator

$$I_2' = \frac{V_{ph}}{\sqrt{\left(r_1 + \frac{r_2' + R'}{2 - s}\right)^2 + (x_1 + x_2')^2}} = \frac{288.67}{\sqrt{(0.13 + 0.836)^2 + (1.2)^2}} = 187.4\ \text{A}$$

and the initial counter current braking (plugging) torque

$$T_{pl} = \left(\frac{3(I_2')^2}{\omega_s}\right)\left(\frac{r_2' + R'}{2 - s}\right) = \frac{3 \times (187.4)^2 \times 0.836}{78.54} = 1121.4\ \text{N} \cdot \text{m}$$

AC braking

The ac braking of an induction motor can be accomplished by disconnecting the machine from the supply and connecting a bank of capacitors to the stator terminals (Fig. 2.31). The magnetizing current of machine is provided by the capacitor bank. The machine operates as an induction generator. The generated energy is dissipated as heat in the rotor circuit bringing the motor to stop. In view of the high cost of capacitors, ac braking is not preferred.

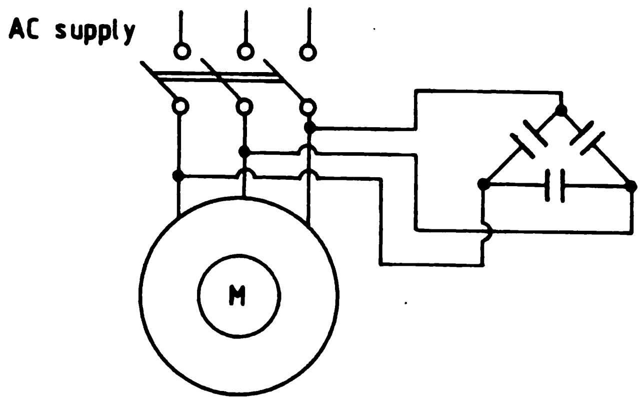


Fig. 2.31 AC braking of an induction motor.

Counter current braking is normally used, if reversal of direction of rotation of the motor is required. Otherwise, dc dynamic braking is usually used.

2.8 SYNCHRONOUS MOTOR

The synchronous motor is used, where constant speed is required. The lagging reactive current drawn by a large number of induction motors used in a factory, is compensated by using an over-excited synchronous motor. The efficiency (92–96%, highest of all motors) and power factor, which can be controlled as required, are very high. If the motor is operated at or near unity power factor, copper loss in the motor along with the size and cost of the inverter needed, is reduced. Also less maintenance is required. Nowadays, all such motors are provided with brushless excitation in which there is no need for a separate dc generator for field excitation. Only a small exciter (ac generator) along with bridge rectifiers is mounted on the rotor shaft, thus, eliminating the slip rings and brushes.

2.8.1 Excitation Control

The synchronous motor can operate at lagging, unity or leading power factor depending upon the field (excitation) current. The expression for the output power developed is

$$P = \omega_m T = \frac{3EV}{x_s} \sin \delta$$

where

V = Input voltage (V) per phase,

E = Induced voltage (V) per phase,

$x_s = 2\pi f L_s$ = Synchronous reactance (Ω) per phase,

L_s = Inductance (H) per phase,

f = Input (line) frequency (Hz),

$\omega_m = 2\pi f\left(\frac{2}{p}\right)$ = Motor speed (synchronous) in rad/s,

p = No. of poles (motor),

P = Output power developed (W),

T = Torque developed (N · m), and

δ = Torque angle = Phase angle between V and E

The motor resistance being small compared to synchronous reactance is neglected. Then, the input power is equal to the output power. It is also assumed that the motor is supplied from an infinite bus, whose voltage and frequency remain constant, in which case the resultant flux in the air gap $\left(\propto \frac{V}{f}\right)$ is constant. The input current is (refer to phasor diagram—Fig. 2.32a)

$$I_a = \frac{V\angle 0° - E\angle -\delta}{x_s}$$

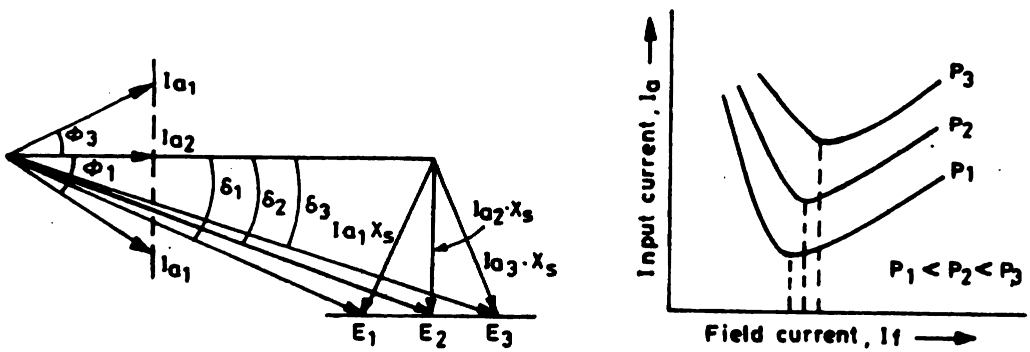
It can also be shown that

$$E \sin \delta = \text{constant}$$

and

$$I_a \cos \theta = \text{constant}$$

The effect of variation of field (excitation) current on input (supply) current for constant output (or input, in this case) power as can be drawn from Fig. 2.32a, is shown in Fig. 2.32b. The graph



(a) Phasor diagram. (b) V-curves.

Fig. 2.32 Synchronous motor delivering constant output power with variable field current.

is in the form of a V-curve. Starting from smaller values of field current (the motor is under excited at that time), when the input power factor is lagging, the power factor goes to unity as the field current is increased. If the field current is increased further leading to overexcitation in the motor, the power factor changes to leading.

2.8.2 Braking of Synchronous Motors

Theoretically, the synchronous motor can be braked, either by counter current braking or by dynamic braking. However, the counter current braking in synchronous motors has the following disadvantages.

1. The synchronous motors are normally of high power ratings. So, heavy disturbances occur with the application of counter current braking to a synchronous motor.

2. The plugging torque is produced by the damper winding. Unless it is specifically designed for this purpose, the counter current braking torque becomes smaller than the dynamic braking torque. In view of the above, dynamic braking is almost always applied for braking the synchronous motors.

Dynamic braking

Dynamic or rheostatic braking of a synchronous motor is performed by disconnecting it from ac mains, and then reconnecting it to a three-phase balanced load. The field excitation remains constant during braking. The machine acts as a generator dissipating its stored energy as heat in the resistive load.

Let M_{af} = mutual inductance between the armature and the field windings,

I_f = excitation current,

ω = instantaneous rotational speed (angular),

R' = braking resistance per phase including the stator resistance.

The voltage (rms) induced per phase,

$$E = \frac{\omega M_{af} I_f}{\sqrt{2}}$$

The braking current,

$$I_b = \frac{\omega M_{af} I_f}{\sqrt{2} \cdot \sqrt{(R')^2 + (\omega L_s)^2}}$$

Since $R' << L_s$, $I_b = \dfrac{M_{af} I_f}{\sqrt{2} L_s} = kI_f$

Therefore, if I_f is kept constant, the braking current also remains constant down to very low speeds. The generated power also remains constant during braking. So, the braking torque increases as the speed decreases. Hence, the torque equation is

$$T_b = \frac{P_b}{\omega} = \frac{3I_b^2 R'}{\omega} = \frac{3\omega(M_{af} I_f)^2}{2[(R')^2 + (\omega L_s)^2]} \tag{2.38}$$

The condition for maximum braking torque is obtained by differentiating Eq. (2.38) with respect to ω and then equating the derivative to zero. The maximum braking torque would occur at the speed, when the machine reactance is equal to the braking resistance.

Example 2.10

The input to an 11,000 V, four-pole, 50 Hz, star connected synchronous motor is 60 A at a power factor of 0.8. The effective resistance and synchronous reactance per phase are 1.0 Ω and 30.0 Ω, respectively. Compute the value of the stator resistance per phase required to be inserted to brake the motor, so that the initial braking current does not exceed 60 A. The field excitation is kept constant at the value, which results in input current of 60 A at a power factor of 0.8 leading. Also, compute the magnitude of the initial braking torque.

Solution

$$r_a = 1\ \Omega, \qquad x_s = 30\ \Omega, \qquad f = 50\ \text{Hz}, \qquad p = 4$$

$$N_s = \frac{120f}{p} = \frac{120 \times 50}{4} = 1500\ \text{rpm}, \qquad \omega_s = \frac{2\pi}{60} N_s = \frac{2\pi \times 1500}{60} = 157.1\ \text{rad/s}$$

Rated voltage per phase $V = \dfrac{11000}{\sqrt{3}} = 6351$ V

Rated current per phase $I = 60$ A

Power factor, $\cos\phi = 0.8$ leading

From Fig. 2.33, we have

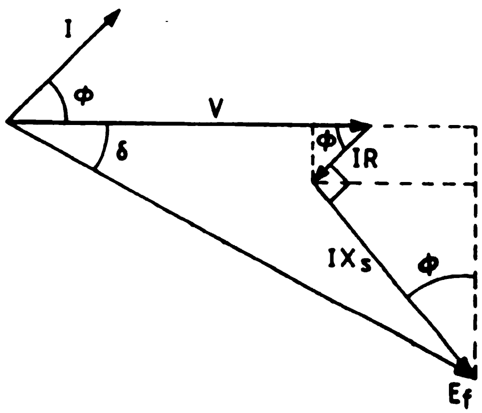


Fig. 2.33 Phasor diagram of the synchronous motor (Example 2.10).

$$E = \sqrt{(V - Ir_a \cos\phi + Ix_s \sin\phi)^2 + (Ir_a \sin\phi + Ix_s \cos\phi)^2}$$

$$= \sqrt{(6351 - 60 \times 1 \times 0.8 + 60 \times 30 \times 0.6)^2 + (60 \times 1 \times 0.6 + 60 \times 30 \times 0.8)^2} = 7529.1\ \text{V}$$

R = stator resistance per phase to be added

$$I_b = \frac{E}{\sqrt{(r_a + R)^2 + (x_s)^2}} = \frac{7529.1}{\sqrt{(1 + R)^2 + (30)^2}} = 60 \text{ A}$$

or

$$R = 120.85 \ \Omega$$

$$\text{Initial braking torque } T_b = \frac{3I_b^2 (r_a + R)}{\omega_s} = \frac{3 \times (60)^2 \times 121.85}{157.1} = 8377 \text{ N} \cdot \text{m}$$

PROBLEMS

2.1 A 220 V, 10 kW, 1200 rpm dc shunt motor has full-load efficiency of 85%. The field resistance and the armature resistance are 110 Ω and 0.25 Ω, respectively. Neglect rotational losses and armature reaction. Calculate the value of the resistance required to be inserted in series with the armature to reduce the speed to 900 rpm, when

(a) the load torque is constant regardless of the speed,

(b) the load torque is directly proportional to the speed, and

(c) the load torque varies as the square of the speed.

2.2 A 400 V, 15 kW, dc shunt motor takes 42 A, and runs at a speed of 1200 rpm. The shunt field has a resistance of 200 Ω. Neglect rotational losses and armature reaction. What value of external resistance must be included in the field circuit, if it is desired to raise the speed to 1500 rpm under the following conditions?

(a) The load torque remains constant at different speeds.

(b) The power remains constant at different speeds.

2.3 A 250 V dc shunt motor has an armature resistance of 0.05 Ω and with rated field excitation has a back emf of 245 V at a speed of 1200 rpm. It is coupled to an overhauling load with a torque of 200 N · m. Determine the lowest speed at which the motor can hold the load by regenerative braking.

2.4 A 500 V dc shunt motor taking an armature current of 240 A, while running at 800 rpm, is braked by disconnecting the armature from the supply and closing it on a resistance of 2.02 Ω, the field excitation remaining constant. The armature has a resistance of 0.5 Ω.

(a) Calculate the initial braking current.

(b) Find the torque at 600 rpm as a percentage of the torque at 800 rpm.

2.5 Construct the natural speed characteristic of a dc series motor having the following ratings:

$$P_{nom} = 3.73 \text{ kW}, \quad V_{nom} = 250 \text{ V}, \quad I_{nom} = 20 \text{ A}, \quad N_{nom} = 1000 \text{ rpm}$$

Determine the value of external resistance required to be inserted in the armature circuit to reduce the speed to 800 rpm and 600 rpm, respectively with armature current, $I_a = I_{nom}$. Construct the rheostatic characteristics according to which the external resistance should be switched for the control.

2.6 Calculate the resistance to be connected across a dc series motor used in a crane, when the supply is cut off, and dynamic braking is applied to limit the speed to 500 rpm, if the descending load exerts a constant load torque of 200 N · m. The magnetization curve of the motor running at 750 rpm is a straight line given by $E = (5.7I_a + 228.6)$ V between $I_a = 30$ A to 50 A. The total motor resistance including the series field resistance is 1.1 Ω.

2.7 A dc series motor has the following parameters:

$$P_{nom} = 15 \text{ kW}, \quad V_{nom} = 220 \text{ V}, \quad r_a = 0.08\ \Omega, \quad r_{se} = 0.17\ \Omega,$$

$$I_{nom} = 80 \text{ A}, \quad N_{nom} = 350 \text{ rpm}$$

What resistance must be connected in series with the armature at the instant of plugging to limit the current to 120 A? Calculate the braking torque at the instant of plugging. Also find the braking torque, when the speed of the motor has fallen to 75% of its rated speed. The magnetization curve of the motor is a straight line given by $E = (1.07I_a + 115)$ V between $I_a = 60$ A to 120 A.

2.8 The following results for a three-phase, 15 kW, 415 V, 50 Hz, six-pole, star-connected induction motor were obtained from the blocked rotor test:

200 V, 50 A, power factor = 0.415

The motor drives a load having a constant torque of 175 N · m. Estimate the possible percentage reduction in the supply voltage before the motor stalls. Assume that the copper losses are equally divided between the stator and the rotor. Neglect magnetizing current.

2.9 What must be the percentage tapping of an auto-transformer starter so that the starting current of an induction motor does not exceed 1.5 times the full-load current? The short-circuit current on normal voltage is 4.5 times the full-load current and the full-load slip is 3%. Calculate the ratio of the starting torque to the full-load torque.

2.10 A 230 V, three-phase, 50 Hz, four-pole, delta-connected, cage induction motor has the following equivalent circuit parameters per phase:

$$r_1 = 0.13\ \Omega, \quad r_2 = 0.32\ \Omega, \quad x_1 = 0.6\ \Omega, \quad x_2 = 1.48\ \Omega, \quad r_m = 250\ \Omega, \quad x_m = 20\ \Omega$$

The full load slip is 5%. The effective stator to rotor turns per phase = 1/1.57. The machine is to be braked from the full-load speed by plugging, when an external resistance of 1.5 Ω per phase in primary terms is inserted into the rotor circuit. Determine the initial plugging torque and the total braking torque, when plugging is applied. Neglect mechanical losses and use the approximate equivalent circuit.

2.11 Dynamic braking is applied with a direct current of 50 A between any two stator terminals of a three-phase, four-pole, delta-connected, cage induction motor running at full-load slip of 5% on 415 V, 50 Hz supply. The standstill rotor impedance of the motor is $(1 + j4)\ \Omega$ per phase referred to the stator and the magnetizing reactance is 28 Ω per phase. Neglect stator impedance and rotational losses. Use the approximate equivalent circuit. Calculate the total braking torque at the instant of application of dynamic braking.

CHAPTER 3

SPEED CONTROL OF DC MOTORS

3.1 BASIC PARAMETERS

Electric motors are used to drive loads of varying characteristics. Precise speed control of electric motors in either direction or their constant speed operation under varying load conditions is required in different applications in industries, electric traction and machine tool, etc. to attain a high rate of production, high quality of products and at the same time to achieve economy in production.

A few examples are cited to illustrate the importance of speed control. Speed control of motors in steel mills is required to a high degree of accuracy in order to avoid sag between the stands. In metal cutting machine tools, the speed of the drive is set and adjusted depending upon the quality of metal to be cut and the tool to be used. In a paper mill, the weight per unit area of paper is determined by the speed of the machine.

The following basic features govern the design of speed control of an electric motor:

Range of speed control. This is expressed as the ratio of $\omega_{max}/\omega_{min} = D$, i.e. 2:1, 4:1, 10:1, etc. For metal cutting machine tools, D ranges from 4:1 to 100:1 and even more. Depending upon the different grades of paper in paper mills, D lies between 3:1 to 20:1.

Smoothness of speed control. It depends upon the number of steps of the starting resistor over the desired range of speed control. The more the number of steps, the smoother would be the speed control. The quality of products depends upon the smoothness of speed control.

Economic justifiability. The method of speed control has to be efficient in terms of cost and energy. Besides these, the reliability and availability of different components and parts need to be taken into account as well.

Stability of operation. It depends upon the hardness of the speed-torque characteristics. If the characteristics are drooping excessively, i.e. having less hardness, the operation tends to be unstable, because drooping characteristics, for a small change in load torque, cause a large change in the speed of the motor.

Direction of speed control. It means whether the speed is controlled below or above the base speed, which is defined as rated speed with rated voltage and full-field excitation. For example, with armature voltage control of a dc shunt motor, the speed can be controlled only in the downward direction with respect to base speed, whereas with field control method, the speed can be controlled only in the upward direction with respect to base speed.

Permissible load at different speeds. Speed control is related to load torque. In a fan or pump drive, the load torque varies as the square of the speed. For speed control of induction motor driven pumps or fans, speed control by voltage variation using ac regulators is the most suitable method. In many applications, speed control is required at constant torque. The drives used in cranes, hoists, rolling mills, paper mills, etc. may be cited as examples. On the other hand, in metal cutting lathes or wheel and axle plants, for example, speed control is required at constant horsepower. Constant linear speed (peripheral speed) and force between the job surface and the cutting tool are the prerequisites for smooth and uniform cutting. As the diameter decreases, the speed increases too. Precise speed control at constant horsepower needs to be achieved.

It is important to determine that the motor capacity is being properly utilized during the complete range of speed control. This can be checked in two ways: by verifying whether the motor is running (a) with rated current, and (b) at constant horsepower throughout the range of speed control. It may be taken into consideration that the cooling conditions deteriorate at low speeds for self-ventilated fan cooled motors.

The constant torque and constant horsepower operations have been illustrated in Fig. 3.1. In dc shunt motors, constant torque operation takes place during speed control below rated speed (base speed) by armature voltage control, while constant horsepower operation takes place during speed control above rated speed (base speed) by excitation control.

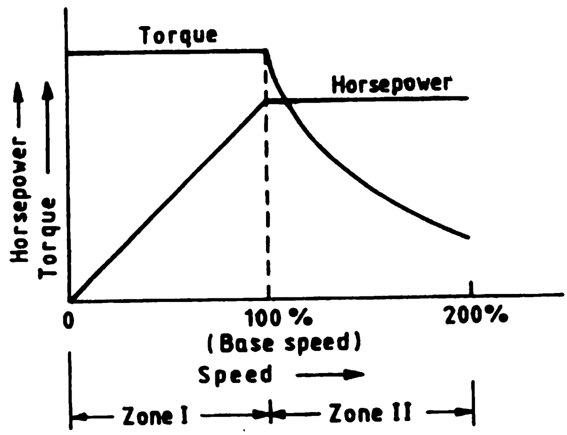


Fig. 3.1 Typical torque-speed and power-speed characteristics of dc shunt motors for the two zones of operation.

The maximum permissible torque will differ with the different methods of speed control. It is the amount of torque that the motor can deliver instantaneously without damage to mechanical parts, commutator and brushes.

3.2 SPEED CONTROL OF DC SHUNT MOTORS

The speed-torque characteristics of dc shunt motors have been described in Chapter 2. The speed equation of a dc motor is given by

$$\omega = \frac{V - I_a(r_a + R_{rh})}{K\phi} \tag{3.1}$$

Hence, the basic speed control methods of a dc motor can be deduced from Eq. (3.1) and these are:

(a) armature voltage control, and

(b) excitation (field current) control.

Standard dc shunt motors are classified as either constant speed or adjustable speed motors. Adjustable speed motors may be operated over a wide range of speed by controlling armature voltage and/or field excitation. The commutation limits the maximum current that can be supplied to a motor such that no sparking occurs. The effect of armature reaction causes certain difficulties. To overcome such effects, adjustable speed motors and nearly constant speed motors are built with a stabilizing series field winding of a few turns around the main poles.

A shunt motor is reversed by interchanging either the armature or the field terminals. If the method of field reversal is adopted, it is likely to cause very high switching voltage surges due to high inductance of the field winding. A thyrite resistor may be used as the field discharge resistor. The speciality of the thyrite resistor is that it allows only a negligible amount of current at normal voltage applied across it. The current flows through a thyrite resistor rather abnormally, if a high voltage appears across its terminals.

3.2.1 Ward-Leonard Method

The speed can be controlled over a wide range in either direction by the Ward-Leonard method. The schematic diagram is shown in Fig. 3.2.

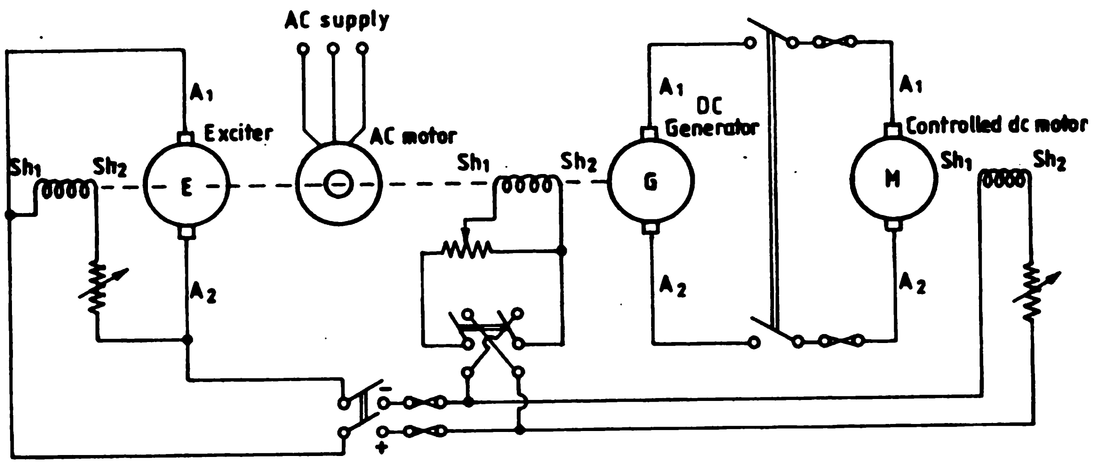


Fig. 3.2 Schematic diagram of the Ward-Leonard method of speed control of dc shunt motors.

The controlled dc motor is supplied from a dc generator. A starter is not needed. The drive motor, which can be an induction or synchronous one, of the motor-generator (M-G) set runs at constant speed. The variable voltage from dc generator is obtained by controlling its field excitation with the help of a potentiometer. Both the dc motor and generator are separately excited. For reversing the dc motor, a double pole double throw (DPDT) switch is used to change the polarity of the field excitation of the dc generator and thereby, the polarity of its dc voltage. The speed control can be effected in two ways:

(1) Constant torque operation by armature voltage control for speeds below base speed.

(2) Constant horsepower operation by field (excitation) control for speeds above base speed

The operating characteristics are shown in Fig. 3.1. The field excitation is kept at maximum value during armature voltage control as it gives maximum torque per unit armature current (ampere). For field excitation control, torque per ampere is reduced to maintain constant horsepower. The range of speed control is normally from 6:1 to 8:1 for motors of medium capacity. The range can be increased to 12:1–16:1 by varying the field excitation of the controlled motor. The upper limit of the speed is restricted by commutation difficulties, whereas the lower limit is restricted by residual magnetism.

It may be noted that, if the field excitation of the dc generator is varied from negative maximum to positive maximum and then back from positive maximum to negative maximum, then the corresponding plot of speed variation with generator field excitation is a hysteresis loop. Full torque is available at any speed. The motor crawls particularly at light loads due to residual magnetism. This crawling is prevented by reducing the effect of the residual magnetism in the following two ways:

(1) By connecting a differential winding across the armature terminals of the generator.

(2) By connecting the field winding of the generator across the armature terminals so as to oppose the residual magnetism. This is called the *suicide connection.*

Advantages

The advantages of the Ward-Leonard drive are:

(a) It has inherent regenerative braking capacity.

(b) A wide range of speed control of the dc motor is obtained in either direction.

(c) The lagging reactive volt-amperes of a plant can be neutralized, by using an over-excited synchronous motor. The overall power factor of the plant also improves.

(d) When the load is intermittent as in rolling mills, the drive motor used is an induction motor with a flywheel mounted on its shaft to smooth out the intermittent loading to a low value. This has been described in Section 5.6. The M-G set provided with a flywheel is known as Ilgner set. The synchronous motor is not suitable for intermittent loading because of its constant speed characteristic.

Disadvantages

The disadvantages of the Ward-Leonard drive are:

(a) High initial cost and low efficiency because of an additional M-G set.

(b) Costly foundation and more floor area are required.

(c) The drive produces noise and requires frequent maintenance.

Nowadays, static Ward-Leonard drives, using two thyristor bridge rectifiers (converters) connected back to back (described in Section 3.4.3), are used to eliminate the above drawbacks.

More precise and smooth speed control over a wide range such as 200:1, and even higher, is accomplished in which there are no sources of noise and vibration. In such an adjustable voltage system, harder torque-speed characteristics can be obtained by providing $i_a r_a$ compensation. Soft starting can also be incorporated.

Non-electrical prime movers can be used to drive the dc generator of the Ward-Leonard drive. For example, in diesel electric locomotives and ship propulsion drives, the generator is driven by a diesel engine or a gas turbine.

Examples of applications are rolling mills, paper mills, mine hoists, etc.

3.2.2 Buck-boost Control

In this method, a variable voltage generator (Buck-boost), driven by a constant speed motor, is connected in series with the dc supply. The schematic connection diagram is given in Fig. 3.3.

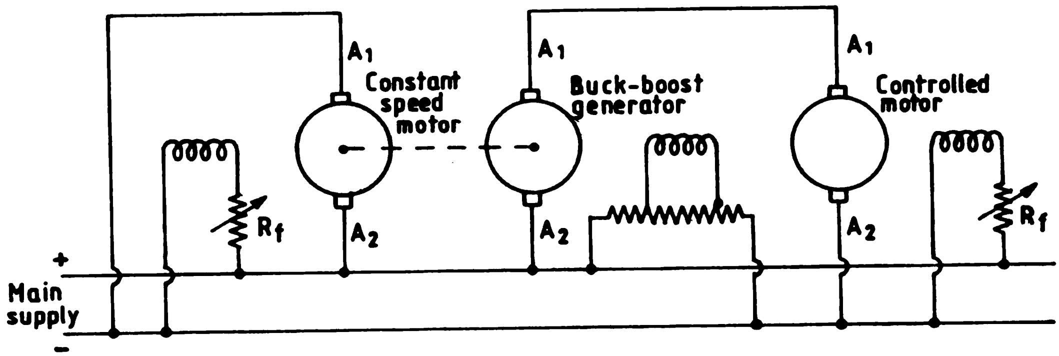


Fig. 3.3 Schematic diagram of Buck-boost method of speed control of dc shunt motors.

At starting, the generator is excited in a direction such that its output voltage is in series buck with the dc supply. The potentiometer is adjusted such that the permissible starting current flows through the armature of the controlled dc motor, whose speed depends upon its field excitation (current) and hence, the output voltage of the generator. As the field current is decreased with the help of the potentiometer, the bucking voltage of the generator decreases too and the net voltage appearing across the motor increases, thus increasing its speed. When the excitation current of the generator is reduced to zero, the motor runs at half the rated (base) speed. Beyond this point (mid-point of the potentiometer), the field current is increased in the opposite direction. The polarity of the generator now reverses, connecting it in series boost with the dc supply. The rated (base) speed is obtained, when the booster voltage is increased by adjustment of its excitation current with the help of the potentiometer such that rated voltage is applied across the motor. The motor field excitation is always at rated value. Beyond the rated (base) speed, the motor speed is controlled by variation of field excitation. The speed control takes place in two modes: (a) armature voltage control, and (b) field excitation (current) control. The relevant performance characteristics are described in Section 3.2.1. Thus, buck-boost control is similar in principle to Ward-Leonard control with the advantages and disadvantages being nearly the same. However, the scheme is economical because of the reduced size of the M-G set.

3.3 SPEED CONTROL OF DC SERIES MOTORS

The speed Eq. (3.1) and the methods of speed control discussed in Section 3.2 are also applicable to dc series motors.

(a) The range of speed control obtained, by inserting resistance in the armature circuit, lies between 2:1 to 3:1 depending upon the load. The hardness of torque-speed characteristics is reduced with the increase in armature resistance. The power loss is high with this method. Series motors are used as crane and traction drives incorporating this method of speed control.

(b) The speed is controlled by variation of the number of field turns, where the speed required is above the base speed.

(c) Armature voltage is varied using the Ward-Leonard method of speed control with two series machines. Alternatively, when a variable voltage source is not available, the series-parallel connection scheme may be employed.

3.3.1 Series-Parallel Control

Two identical series motors, mechanically coupled, are connected in series and parallel as shown in Fig. 3.4 for the purpose of speed control. The capacity of each motor is approximately half of the capacity of the load.

Several motors may also be connected in series-parallel. The size of each motor as well as the moment of inertia is then decreased, thus shortening the time of starting and braking. Two steps of speed are provided without waste of power in the series resistance. If intermediate speeds are required, a series resistance is inserted in the armature circuit. For full utilization of motor capacity, speed control should be done at constant output torque. Such drives are used in ladle cranes, electric trains, trolley cars, blast furnace skip hoists, etc.

3.3.2 Speed Control (during Lowering) of a DC Series Motor in a Crane using Dynamic Braking

Emergency braking

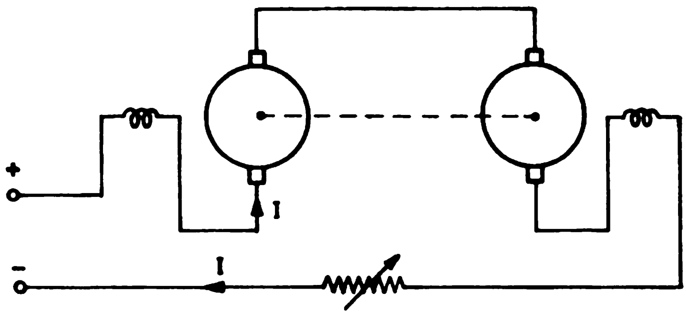
Dynamic braking may be used to control the speed of a hoist motor in a crane, which is a dc series motor. It is assumed that there is a load on the crane hook. As an initial approach, let us consider what happens if power failure occurs, while the crane is in operation.

Immediately upon occurrence of power failure, the armature is connected in series with the field and a suitable external resistance. The motor will be overhauled by the load on the hook due to gravity and act as a self-excited series generator. Care should be taken to connect the series field winding such that the residual and series fields are in the same direction.

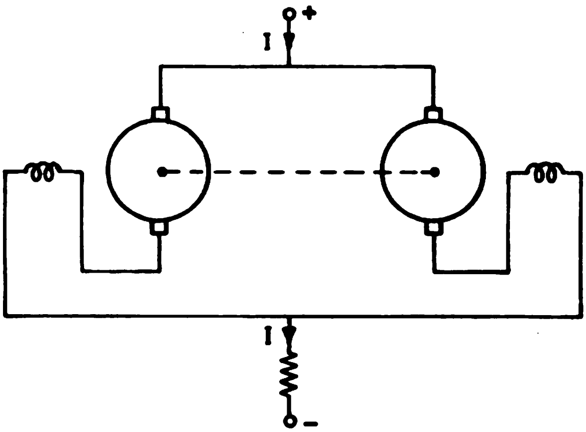
The emf and current relation is given by

$$-E = I_a(R_{\text{int}} + R_{\text{ext}}) \tag{3.2}$$

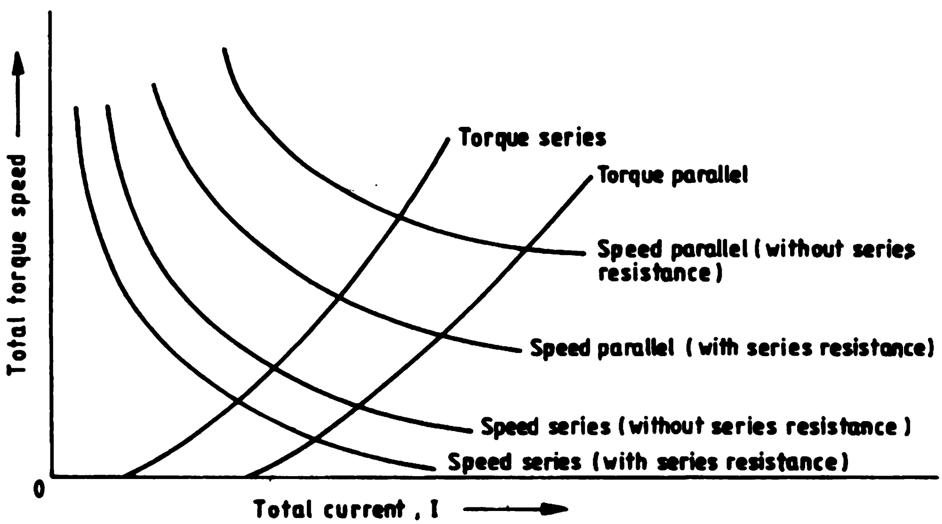
The emf is negative, because the motor now runs in the reverse direction. Earlier, it was rotating in the forward direction as motor during hoisting. Now, the armature terminals have reversed their polarity. Dynamic braking torque is produced, while the motor is being driven by the overhauling load. The speed-torque characteristics for emergency braking are shown in Fig. 3.5. These characteristics depend upon the external resistance and lie in the fourth quadrant.



(a) Series connection.



(b) Parallel connection.



(c) Total characteristics of a pair of series motors.

Fig. 3.4 Series-parallel control of a pair of dc series motors.

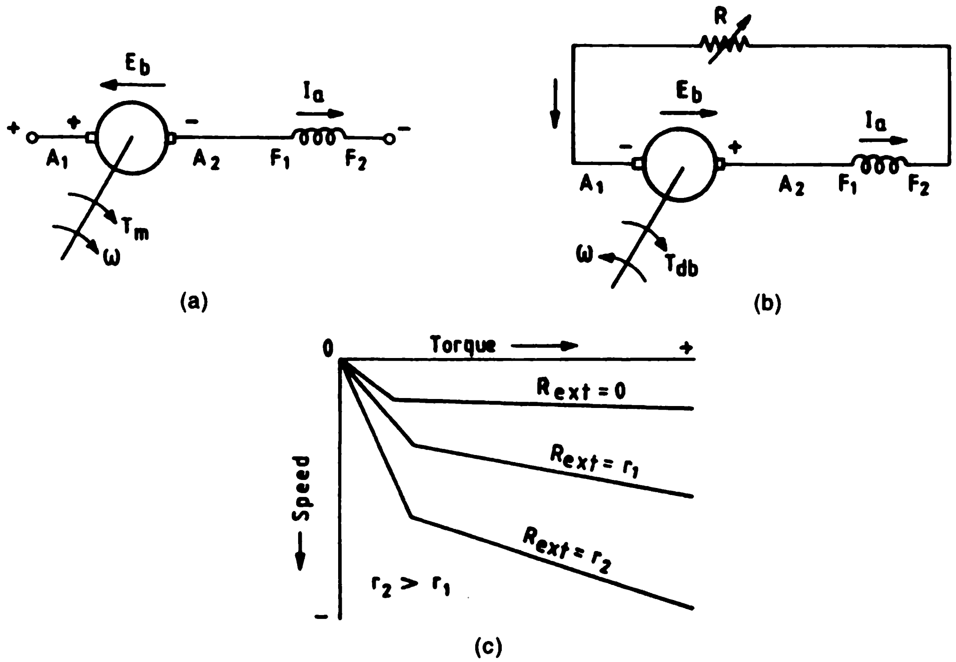


Fig. 3.5 Circuits and speed-torque characteristics of dc series motors used in cranes: (a) Circuit for motor (hoist) operation; (b) Circuit for dynamic (emergency) braking during lowering; (c) Speed-torque characteristics for dynamic braking.

Dynamic braking during lowering

The lowering speed can be controlled suitably, if the armature and field circuits are connected in parallel as shown in Fig. 3.6 instead of in series. The series motor acts as a separately excited generator, while the load on the hook overhauls it under the influence of gravity. E is negative under this condition. The current will flow in the same direction as during motoring. Dynamic braking torque is produced.

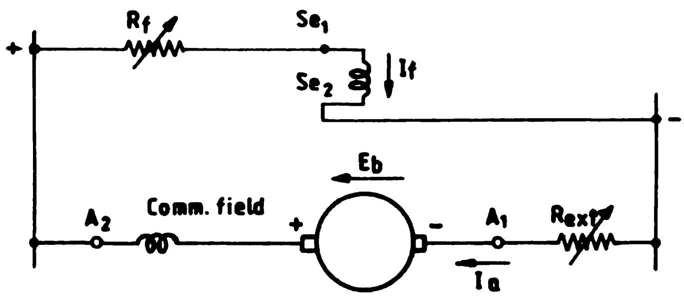


Fig. 3.6 Circuit connections of a dc series motor in a crane for lowering using dynamic braking.

The armature current-voltage relationship can be expressed as

$$-E = V + I_a (R_{\text{int}} + R_{\text{ext}}) \tag{3.3}$$

The speed and torque are obtained from E and I_a, respectively. The voltage E also depends upon field current I_f. Lowering control can therefore be done independently by variation of field current I_f with the help of R_f and/or R_{ext}. Figure 3.7 shows the speed-torque characteristics for different field currents keeping $R_{ext} = 0$.

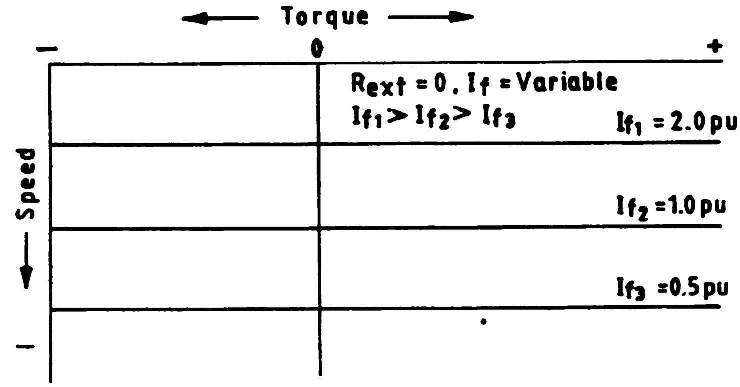


Fig. 3.7 Speed-torque characteristics of the dc series motor of Fig. 3.6 for various values of field current with $R_{ext} = 0$.

These characteristics are hard. The lowering speed of the motor increases with decrease of field current. These characteristics are not satisfactory for lowering a load at a slow speed, as a very high value of field current is required. These characteristics are also not suitable for accelerating and retarding a load because of their hardness. These problems can be overcome either by using a resistance in series with the armature or by using both armature and field resistances. The speed-torque characteristics are, in either case, straight lines according to Eq. (3.2). Excessive weakening of the field to achieve high lowering speeds, gives rise to instability, because armature reaction predominates. The speed-torque characteristics for different values of the resistance in the armature circuit keeping the field circuit resistance fixed are shown in Fig. 3.8.

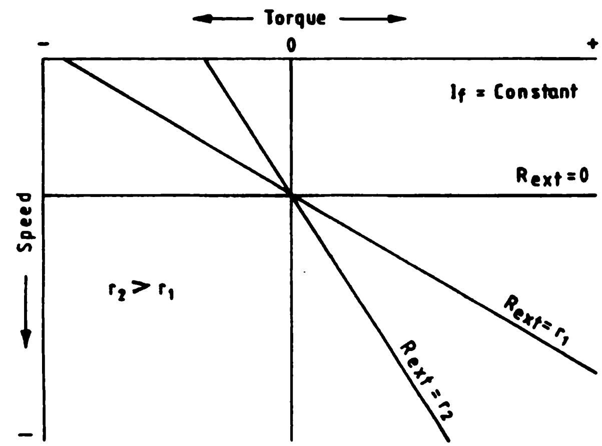


Fig. 3.8 Speed-torque characteristics of the dc series motor of Fig. 3.6 for different values of the armature circuit resistance.

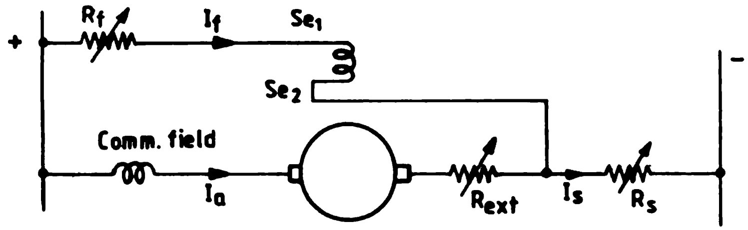


Fig. 3.9(a) Circuit of a dc series motor used in cranes for lowering during dynamic braking using R_{ext}, R_f and R_s.

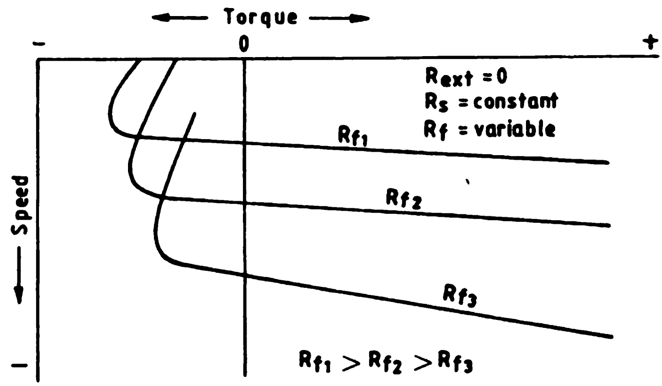


Fig. 3.9(b) Speed-torque characteristics of the dc series motor of Fig 3.9(a) for various values of field circuit resistance.

The armature resistance control for dynamic braking during lowering is not suitable for constant speed operation. The speed-torque characteristics are further modified by connecting a third rheostat in series with both the field and armature circuits as shown in Fig. 3.9a. This type of connection has considerable applications in dynamic braking lowering control of crane hoists. The speed-torque characteristics are shown in Fig. 3.9b.

Example 3.1

A 400 V dc shunt motor takes 110 A and runs at 900 rpm and provides 35 kW of power to a mechanical load in the rated condition. The armature and field resistances of the motor are 0.18 Ω and 140 Ω, respectively. Compare the efficiencies of the motor at 500 rpm at rated torque and 1500 rpm at rated power.

Solution

$V = 400$ V, $I = 110$ A, $R_f = 140\ \Omega$, $R_a = 0.18\ \Omega$, $N_1 = 900$ rpm

At rated condition

Output power $P_{o1} = 35$ kW

Input power $P_{i1} = VI = 400 \times 110 = 44$ kW

$$\text{Efficiency}\,\eta = \frac{P_{o1}}{P_{i1}} = \frac{35}{44} = 0.7955 = 79.55\%$$

Field current $I_{f1} = \dfrac{V}{R_f} = \dfrac{400}{140} = 2.857$ A

Armature current $I_a = I - I_{f1} = 110 - 2.857 = 107.143$ A

Armature copper loss = $I_a^2 R_a = (107.143)^2 \times 0.18 = 2.066$ kW

Field copper loss = $VI_{f1} = 400 \times 2.857 = 1.143$ kW

Constant (core and rotational) losses = $P_i - P_o - I_a^2 R_a - VI_{f1}$

$= 44 - 35 - 2.066 - 1.143 = 5.791$ kW

Angular speed $\omega_1 = \dfrac{2\pi N_1}{60} = \dfrac{2\pi \times 900}{60} = 94.25$ rad/s

Shaft (rated) torque $T_1 = \dfrac{P_{o1}}{\omega_1} = \dfrac{35 \times 10^3}{94.25} = 371.35$ N · m

(a) At $N_2 = 500$ rpm with rated torque

Angular speed $\omega_2 = \dfrac{2\pi \times 500}{60} = 52.36$ rad/s

Output power $P_{o2} = \omega_2 T_1 = 52.36 \times 371.35 = 19.444$ kW

Armature current I_a remains constant, as flux ϕ_1 or field current I_{f1} is constant.
Input power P_{i2}

= Output power + Armature copper loss + Field copper loss + Constant losses

= 19.444 + 2.066 + 1.143 + 5.791 = 28.444 kW

Efficiency $\eta = \dfrac{19.444}{28.444} = 0.6836 = 68.36\%$

(b) At $N_3 = 1500$ rpm with rated power

Output power $P_{o3} = 35.0$ kW

Armature current I_a remains constant.

Neglecting saturation, $\phi \propto I_f \propto \dfrac{1}{N}$, $I_{f3} = \dfrac{N_1}{N_3} I_{f1} = \dfrac{900}{1500} \times 2.857 = 1.714$ A

Field copper loss = $VI_{f3} = 400 \times 1.714 = 0.686$ kW

Input power $P_{i3} = 35.0 + 2.066 + 0.686 + 5.791 = 43.543$ kW

Efficiency $\eta = \dfrac{35.0}{43.543} = 0.8038 = 80.38\%$

Example 3.2

A dc series motor (Fig. 3.10) working off a 500 V system runs at a speed of 800 rpm, when the load is such that the current is 120 A. The resistance of the motor is 0.15 Ω, of which 0.04 Ω is the series field resistance. Assuming that the flux is proportional to the armature current, calculate the speed,

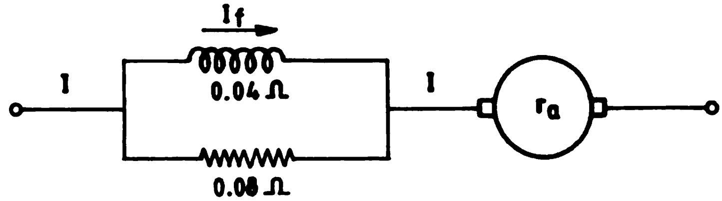


Fig. 3.10 Circuit connections of the dc series motor with the series field diverter (Example 3.2).

(a) when the armature current is 60 A and the supply voltage is 550 V.

(b) when the torque is one-half of the full-load value and the field resistance is connected in parallel with a diverting resistance of 0.08 Ω, the supply voltage being 500 V.

(c) when a resistance of 1.0 Ω is connected in series with the motor and the torque is 75% of the full load value, the supply voltage being constant at 500 V.

Solution

$r_m = 0.15\ \Omega,\quad r_{se} = 0.04\ \Omega,\quad r_a = r_m - r_{se} = 0.15 - 0.04 = 0.11\ \Omega,$

$V_1 = 500\ \text{V},\quad I_1 = 120\ \text{A},\quad V_2 = 550\ \text{V},\quad I_2 = 60\ \text{A},\quad N_1 = 800\ \text{rpm}$

(a) $E_{b1} = V_1 - I_1 r_m = 500 - 120 \times 0.15 = 482$ V

$E_{b2} = 550 - 60 \times 0.15 = 541$ V

Now

$$\frac{N_2}{N_1} = \frac{E_{b2}}{E_{b1}} \times \frac{\phi_1}{\phi_2} = \frac{E_{b2}}{E_{b1}} \times \frac{I_{se1}}{I_{se2}},\ \text{as } \phi \propto I_{se}(I_{se} = I_a = I)$$

Therefore

$$N_2 = \frac{E_{b2}\, I_{se1}}{E_{b1}\, I_{se2}} N_1 = \frac{541 \times 120 \times 800}{482 \times 60} = 1796\ \text{rpm}$$

(b) $r_d = 0.08\ \Omega$

$$I_{se3} = \frac{r_d}{r_{se} + r_d} I_3 = \frac{0.08}{0.04 + 0.08} I_3 = 0.6667 I_3$$

$$T \propto \phi I$$

Therefore

$$T_3 = K\phi_3\, I_3 = 0.6667\ K(I_3)^2 \quad (\because\ \phi_3 \propto I_{se3})$$

Now

$$T_3 = \frac{1}{2}T_1$$

Therefore

$$\frac{T_3}{T_1} = \frac{0.6667 \times (I_3)^2}{(I_1)^2}$$

or

$$\frac{1}{2} = \frac{0.6667 \times (I_3)^2}{(120)^2}$$

Therefore

$$I_3 = \frac{120}{\sqrt{2 \times 0.6667}} = 103.9 \text{ A}$$

$$I_{se3} = 0.6667 \times 103.9 = 69.27 \text{ A}$$

$$E_{b3} = V_1 - I_3 r_a - I_{se3} r_{se} = 500 - (103.9 \times 0.11) - (69.27 \times 0.04) = 485.8 \text{ V}$$

$$N_3 = \frac{E_{b3}\, I_{se1}}{E_{b1}\, I_{se3}} N_1 = \frac{485.8 \times 120 \times 800}{482 \times 69.27} = 1397 \text{ rpm}$$

(c) $T_4 = 0.75T_1$

or

$$\frac{T_4}{T_1} = \frac{I_4^2}{I_1^2}$$

Therefore

$$I_4^2 = 0.75 \times (120)^2$$

or

$$I_4 = 103.9 \text{ A}$$

$$R = r_{se} + r_a + R_{rh} = 0.04 + 0.11 + 1.0 = 1.15\ \Omega$$

Therefore

$$E_{b4} = V_1 - I_4 R = 500 - (103.9 \times 1.15) = 380.5 \text{ V}$$

Hence

$$N_4 = \frac{E_{b4}\, I_{se1}}{E_{b1}\, I_{se4}} N_1 = \frac{380.5 \times 120 \times 800}{482 \times 103.9} = 729 \text{ rpm}$$

Example 3.3

Two dc series motors with different air gaps, but otherwise identical, run at 700 and 750 rpm, respectively, when taking 50 A at 500 V. The total resistance of each motor is 0.36 Ω. If the motors are mechanically coupled, and connected in series to a 500 V supply and take a current of 50 A, calculate:

(a) the speed at which the set will run, and

(b) the voltage across each machine.

Solution

$V = 500$ V, $\quad I = 50$ A, $\quad r_m = 0.36\ \Omega$, $\quad N_1 = 700$ rpm, $\quad N_2 = 750$ rpm

$$\omega_1 = \frac{2\pi N_1}{60} = \frac{2\pi \times 700}{60} = 73.3 \text{ rad/s}, \qquad \omega_2 = \frac{2\pi \times 750}{60} = 78.54 \text{ rad/s}$$

When the motors are running independently, we have

$$E_{b1} = E_{b2} = V - I_a r_m = 500 - (50 \times 0.36) = 500 - 18 = 482 \text{ V}$$

$$K\phi_1 = \frac{E_{b1}}{\omega_1} = \frac{482}{73.3} = 6.576 \text{ V} \cdot \text{s/rad}$$

$$K\phi_2 = \frac{482}{78.54} = 6.137 \text{ V} \cdot \text{s/rad}$$

Now

$$E_b = K\phi\, \omega$$

(a) Assume that the set will run at the speed N_3, when mechanically coupled.

The voltage across machine 1, $V_1 = E_{b1} + I_a r_m = 6.576\ \omega_3 + 18$

The voltage across machine 2, $V_2 = 6.137\ \omega_3 + 18$

Now

$$V = V_1 + V_2$$

Therefore

$$(6.576 + 6.137)\omega_3 + 36 = 500$$

or

$$\omega_3 = \frac{500 - 36}{6.576 + 6.137} = 36.5 \text{ rad/s}$$

Therefore

$$N_3 = \frac{60\omega_3}{2\pi} = \frac{60 \times 36.5}{2\pi} = 349 \text{ rpm}$$

(b)

$$V_1 = (6.576 \times 36.5) + 18 = 258.0 \text{ V}$$

$$V_2 = V - V_1 = 500 - 258.0 = 242.0 \text{ V}$$

Example 3.4

A 240 V dc series motor has an armature resistance of 0.42 Ω and a series field resistance of 0.18 Ω.

The speed is 500 rpm, when the current is 36 A. What will be the motor speed, when the load reduces the line current to 21 A? Assume 2 V brush drop, and also that flux is proportional to current.

Solution

$V = 240$ V, $r_a = 0.42\ \Omega$, $r_{se} = 0.18\ \Omega$, $r_m = r_a + r_{se} = 0.42 + 0.18 = 0.6\ \Omega$,

$I_{a1} = 36$ A, $I_{a2} = 21$ A, $N_1 = 500$ rpm

Now

$$E_b = V - I_a r_m - 2.0$$

Therefore

$$E_{b1} = 240 - (36 \times 0.6) - 2 = 216.4 \text{ V}$$

and

$$E_{b2} = 240 - (21 \times 0.6) - 2 = 225.4 \text{ V}$$

$$\frac{N_2}{N_1} = \frac{E_{b2}\,\phi_1}{E_{b1}\phi_2} = \frac{E_{b2}\,I_{se1}}{E_{b1}\,I_{se2}}, \text{ as } \phi \propto I_{se}\ (I_{se} = I_a)$$

Therefore

$$N_2 = \frac{E_{b2}\,I_{se1}}{E_{b1}\,I_{se2}} N_1 = \frac{225.4 \times 36 \times 500}{216.4 \times 21} = 893 \text{ rpm}$$

3.4 SOLID STATE DC DRIVES

In the last thirty years, the development of various solid state switching devices in the thyristor/transistor families along with varieties of analog/digital chips used in control/firing circuits, has made an impact in the area of dc drives. These power electronic (solid state) controllers are of two types:

(1) Thyristor bridge rectifiers (converters) supplied from ac supply.

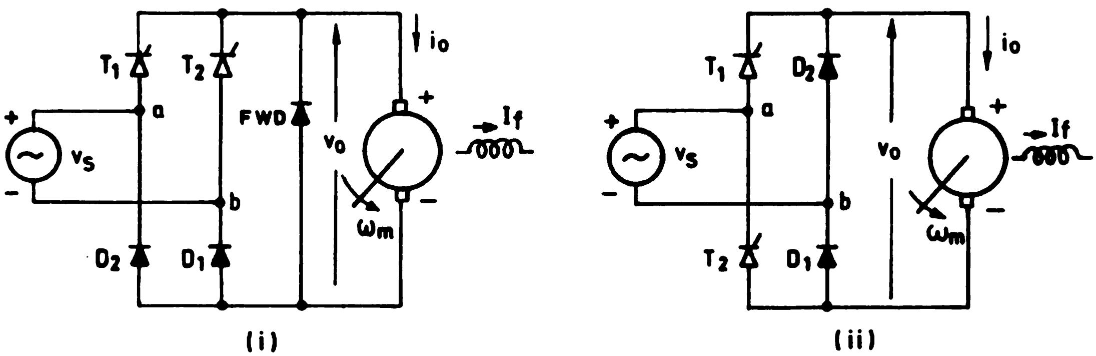
(2) Choppers fed from dc supply.

The first type is used for speed control of dc shunt motors, whereas the second type is mainly used for speed control of dc series motors. In the second type, with the recent advances in the ratings of new types of devices, such as Gate Turn-Off (GTO) thyristors, which can be turned off by pulses fed at the gate terminal, as contrasted with thyristors which need commutation circuits for turning off, the choppers have become less bulky and more reliable. The development of power transistors and allied devices of higher ratings has made the design of choppers simpler as these can be switched on and off by a proper base signal. The use of these devices with simpler commutation circuits has also ushered in changes in the rectifier (converter) circuits, where the demerits of phase control have been eliminated and new circuits with symmetrical firing for improved power factor are being increasingly employed. The cost of these drive systems has reduced and reliability enhanced as the control schemes including firing circuits have been simplified by making use of analog/digital chips and/or microprocessor-based systems.

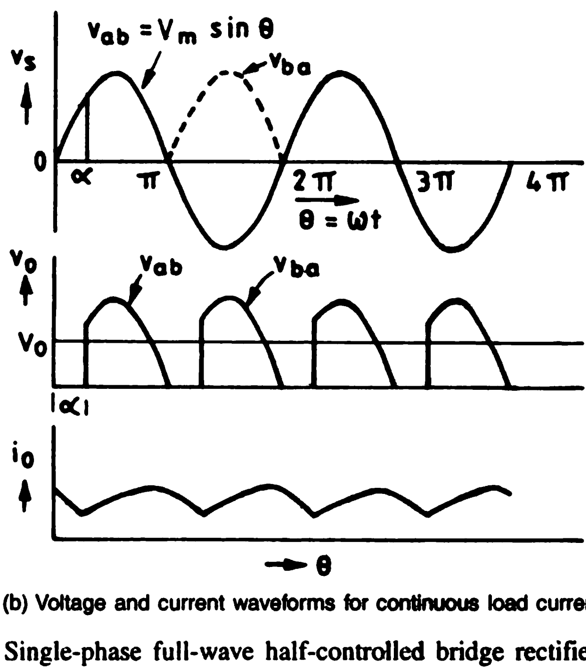
3.4.1 Thyristor Bridge Rectifier Circuits

Single-phase half-controlled bridge rectifier

The single-phase full-wave half-controlled bridge rectifier (converter) with dc motor load is shown in Fig. 3.11a. The first type (Fig. 3.11a(i)) uses two thyristors and three diodes, one of them being a freewheeling (FWD) one, while the second type (Fig. 3.11a(ii)) has two thyristors and two diodes, to be also used in freewheeling mode. This type is suitable for one quadrant operation, with the output voltage being positive and the current in the thyristor unidirectional. The thyristors are fired once in each cycle, T_1 in positive half and T_2 in negative with delay angle, α. The voltage and current waveforms with dc motor load are shown in Fig. 3.11b. The current is continuous for low values of firing delay angle α and high load torque, while the current is discontinuous for large



(a) Power circuits with dc motor load.



(b) Voltage and current waveforms for continuous load current.

Fig. 3.11 Single-phase full-wave half-controlled bridge rectifier (converter).

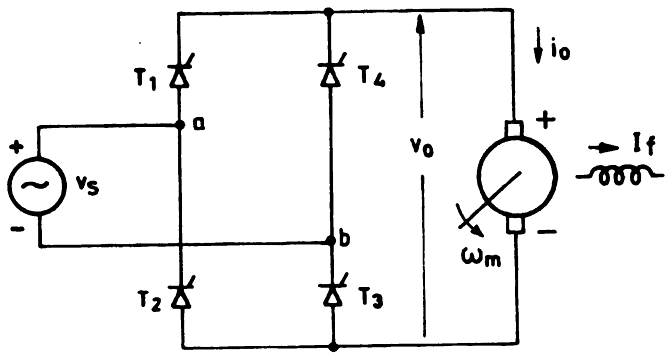
values of delay angle α and low load torque. The thyristor T_1 conducts from α to π, and then the freewheeling action takes place from π to $(\pi + \alpha)$. This is repeated for the negative half with T_2 in place of T_1. The output voltage with continuous current is

$$V_{dc} = 0.9\ V\,(1 + \cos\alpha) \qquad \text{for } \pi > \alpha > 0 \tag{3.4}$$

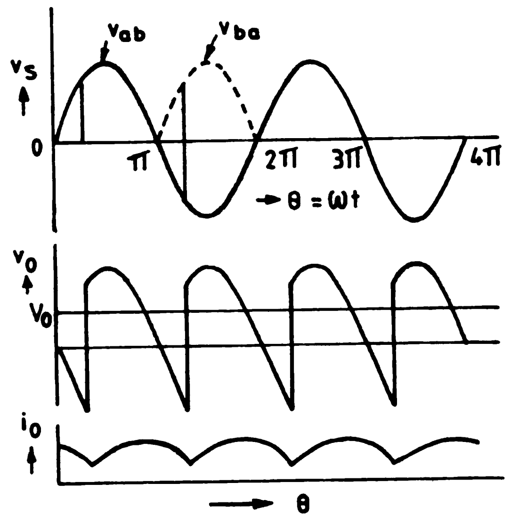
Single-phase full-controlled bridge rectifier

The single-phase full-controlled bridge rectifier with four thyristors is shown in Fig. 3.12a along with continuous voltage and current waveform for motoring action (Fig. 3.12b). This circuit is suitable for two-quadrant operation as the output voltage can be both positive and negative, the current in the thyristors being unidirectional. The output voltage with continuous current is

$$V_{dc} = 0.9\ V \cos\alpha \qquad \text{for } \pi > \alpha > 0 \tag{3.5}$$



(a) Power circuits with dc motor load.



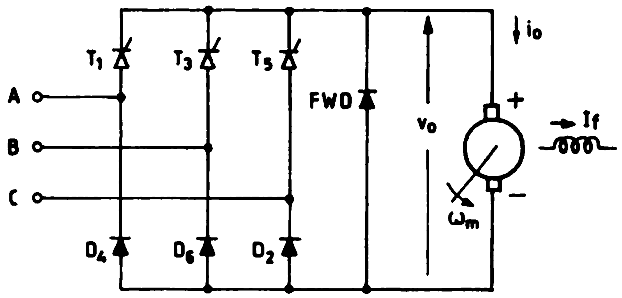
(b) Voltage and current waveforms for continuous load current.

Fig. 3.12 Single-phase full-controlled bridge rectifier.

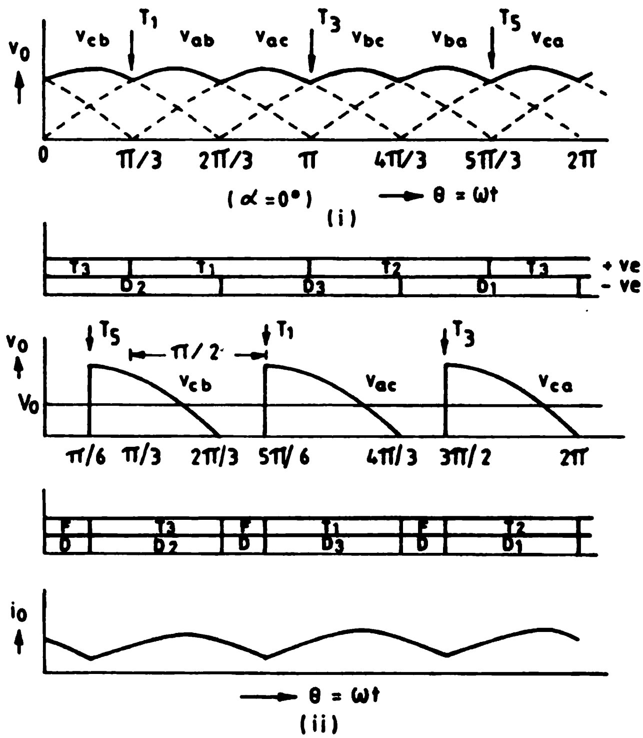
For $\alpha < \pi/2$, the output voltage is positive (motoring action in the first quadrant) while for $\alpha > \pi/2$ the output is negative (regenerative braking in the fourth quadrant). In most cases, the current will be discontinuous both in motoring and braking modes, especially with large delay angle α. The single-phase bridge rectifier is mostly used for low power drives.

Three-phase half-controlled bridge rectifier

A three-phase half-controlled bridge rectifier with three thyristors and three diodes is shown in Fig. 3.13a along with the voltage and current (continuous) waveforms (Fig. 3.13b). The action is



(a) Power circuit with dc motor load.



(b) Voltages: (i) $\alpha = 0°$, (ii) $\alpha = 90°$ and current ($\alpha = 90°$) waveforms for continuous load current.

Fig. 3.13 Three-phase half-controlled bridge rectifier.

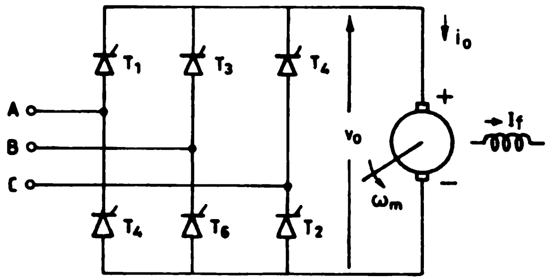
similar to the case of the single-phase type. The thyristors are fired at delay angle α from the natural commutation, while the diodes are naturally commutated. The voltage waveform is continuous for $\alpha < \pi/3$ and discontinuous for $\alpha > \pi/3$. The output voltage (with continuous current) is

$$V_{dc} = 1.35\ V_L(1 + \cos \alpha) \quad \text{for } \pi > \alpha > 0 \tag{3.6}$$

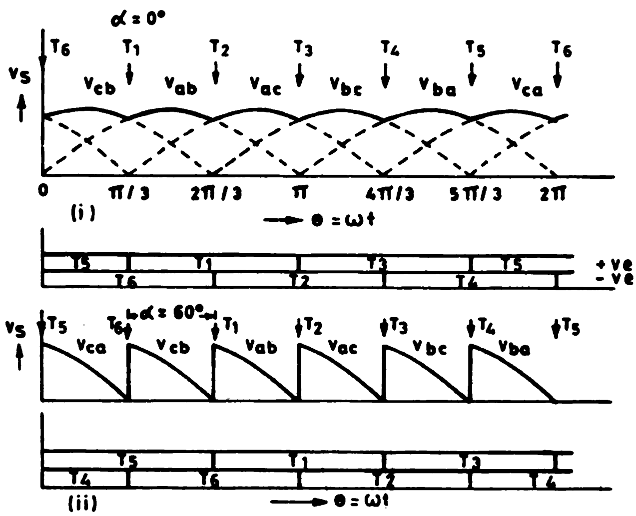
Figure 3.13b(i) shows the voltage waveform for the diode bridge or thyristor bridge for $\alpha = 0°$, where the thyristor T_1 is fired at $\theta = 60°$ with thyristor T_1 and diode D_6 conducting (T_5 going off) for 60°, when commutation occurs from D_6 to D_2. At $\theta = 180°$, the thyristor, T_3 is fired with T_1 going off. If $\alpha = 90°$, T_1 is fired at $\theta = 150°$, with T_1 and D_2 conducting for $90°(\theta = 240°)$, when freewheeling diode (FD) takes over, as the instantaneous voltage goes negative. T_3 is fired at 270° with T_3 and D_4 conducting. This is shown in Fig. 3.13b(ii).

Three-phase full-controlled bridge rectifier

A three-phase full-controlled bridge rectifier with six thyristors is shown in Fig. 3.14a along with waveforms (Fig. 3.14b). The thyristors are fired in sequence with delay angle α, with each thyristor



(a) Power circuits with dc motor load.



(b) Voltages: (i) $\alpha = 0°$, (ii) $\alpha = 60°$ and current ($\alpha = 60°$) waveforms for continuous load current.

Fig. 3.14 Three-phase full-controlled bridge rectifier.

conducting for angle 120°. Two thyristors conduct at a time. When thyristor, T_1 is triggered, the thyristors T_1 and T_6 start conducting, with thyristor T_5 going off. Prior to this, the thyristors T_5 and T_6 are conducting. When thyristor T_2 is triggered after a delay of 60° from the instant T_1 is triggered, the thyristors, T_1 and T_2 start conducting, with thyristor, T_6 going off. This sequence is repeated at an interval of 60°. The output voltage is positive for $\alpha < \pi/2$, and negative for $\alpha > \pi/2$. The output voltage with continuous current is

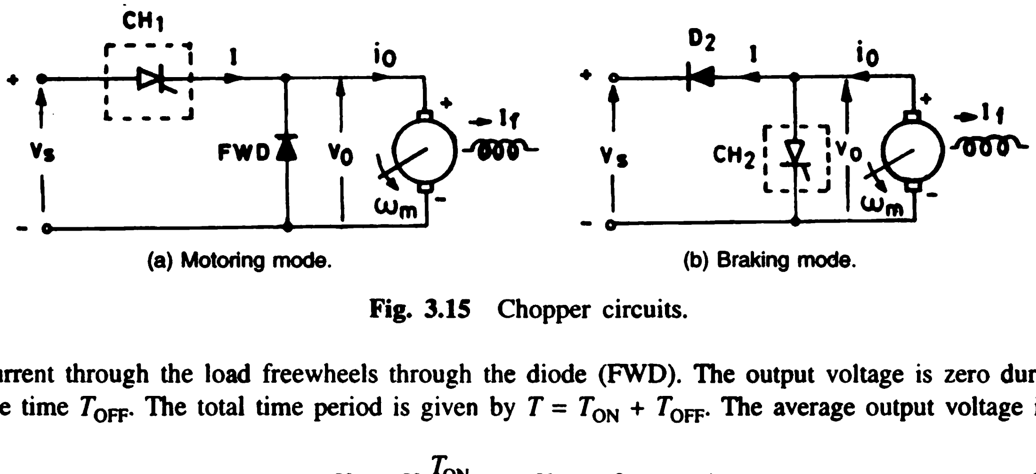
$$V_{dc} = 1.35\ V_L \cos\alpha \quad \text{for } \pi > \alpha > 0 \tag{3.7}$$

The current is mostly continuous, except for large delay in motoring mode, but can be discontinuous in braking mode. In all cases, these circuits are used for armature voltage control to change the speed of dc shunt motors, while the field is fed from a diode bridge. These bridge circuits are used for high power drives. As the thyristors conduct for about one-third of the cycle, the average current in the thyristors is less than rated motor current.

3.4.2 Chopper Circuits

Step-down chopper

The simple chopper can be explained by means of a switch as shown in Fig. 3.15a. When the switch is ON, the output voltage appears across the load. If the switch is made OFF after time T_{ON}, the



(a) Motoring mode. (b) Braking mode.

Fig. 3.15 Chopper circuits.

current through the load freewheels through the diode (FWD). The output voltage is zero during the time T_{OFF}. The total time period is given by $T = T_{ON} + T_{OFF}$. The average output voltage is

$$V_{dc} = V_s \frac{T_{ON}}{T} = \alpha V_s \qquad \text{for } \alpha < 1 \tag{3.8}$$

Thus, by keeping time period T or chopping frequency $f = 1/T$ constant and varying T_{ON} ($T > T_{ON} > 0$), the average output voltage can be varied. The T_{ON} period can be varied from 10% to 90% of time period T. This type of chopper is also called the step-down chopper, as $V_{dc} < V_s$. In this case, the load has been assumed to be inductive with or without any back emf. In certain cases, T_{OFF} is kept constant and only T_{ON} is varied, thus varying time period T or chopping frequency f. The chopper switch, in earlier days, used a thyristor, which required an additional thyristor to switch it off. Nowadays, a GTO is used, which can be turned off through its gate. Also, a power transistor or power MOSFET can be used instead. The chopping frequency is increased, when a power transistor/MOSFET is used.

One-quadrant chopper-fed drive

The chopper can be utilized for speed control, when it is used in the armature circuit of a dc separately excited motor, the field current being kept constant. The variable voltage output obtained is used to vary the motor speed. The speed is proportional to the output voltage, which is controlled by ON time T_{ON}, the time period T being kept constant in this case. The field is supplied from a transistor chopper with the diode (FWD) connected across the field. The high inductance of the field produces continuous current in the field circuit. This is only suitable for one-quadrant (quadrant I) operation as the output voltage is positive and with the current through the motor being also positive, the motoring torque produced is positive and the motor rotates in the forward direction. This chopper can also be used for speed control of dc series motors. For operation in quadrant III (reverse motoring), this chopper can be used too, but must be connected in the reverse direction to the armature with the direction of field current remaining the same. For the current in the armature circuit to remain continuous, especially during the OFF period, the inductance of the armature must be sufficient and the chopping frequency of the chopper high. The chopping frequency of transistor choppers, mostly being used nowadays, is usually high. The analysis follows the same method as described in Section 3.4.3.

Step-up chopper

To use a chopper for braking, the circuit of Fig. 3.15a is modified as shown in Fig. 3.15b. The positions of switch and diode are interchanged here. Now, with the motor running as a generator, the chopper is put on for time, T_{ON}. As the generator is short-circuited, the current builds up in the armature inductance. Now, when the chopper is put off, after time T_{ON}, the current in the armature inductance tends to decrease, thus reversing the voltage developed in it. As the current tends to flow in the same direction, the generator feeds current into the voltage source. Therefore, the speed of the motor now, acting as a generator, decreases, thus giving rise to regenerative braking. It can be shown that the output voltage across the load is

$$V_{dc} = V_s \frac{T}{T - T_{ON}} = \frac{V_s}{1-\alpha} \qquad \text{for } \alpha < 1 \tag{3.9}$$

As $\alpha \left(= \frac{T_{ON}}{T} \right) < 1$, $V_{dc} > V_s$. This chopper is, therefore, called the step-up chopper. It is used for operation in the second quadrant. By combining it with a step-down chopper, two/four-quadrant operation is possible.

Thyristor chopper

The circuit configuration of a thyristor chopper is shown in Fig. 3.16. With the auxiliary thyristor T_a switched on, the capacitor C will be charged with the point E becoming positive and thus the thyristor T_a going off. Then, after some time, when the main thyristor T_m is switched on, the voltage across the capacitor C is reversed with the point F becoming positive, as current through C flows only for half cycle, blocked by the diode as the current tends to reverse. Thus, the capacitor C remains charged for the main thyristor T_m to be put off. As the auxiliary thyristor T_a is put on, the reverse voltage is switched across T_m, thus putting it off. Then the capacitor is charged in the opposite direction, as stated earlier. Thus, an auxiliary thyristor along with additional components is needed for a thyristor chopper.

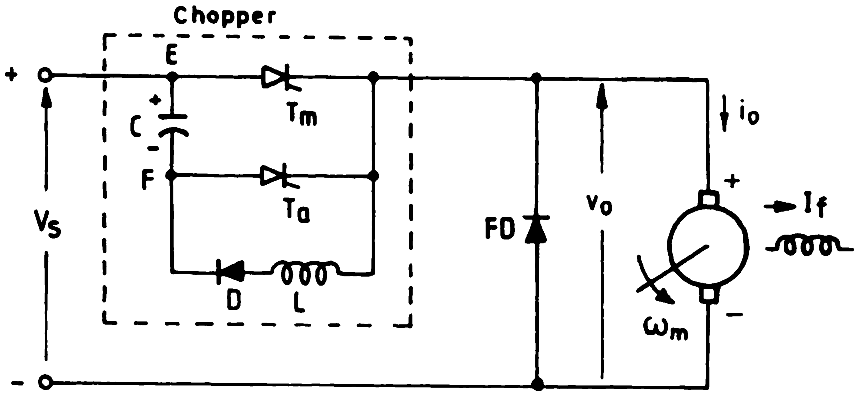


Fig. 3.16 Thyristor chopper circuit (voltage commutation).

Two/four-quadrant chopper-fed drive

In earlier subsections, only one-quadrant operation, either in quadrant I or in quadrant II was presented. By combining these, different types of operations can be performed. The transistorized chopper drive (Fig. 3.17) is suitable for two-quadrant operation. Transistor Q_1 and diode D_2 as (FWD) function as a chopper circuit for operation in quadrant I. Similarly, transistor Q_2 and diode D_1 act as a chopper circuit for operation in quadrant II.

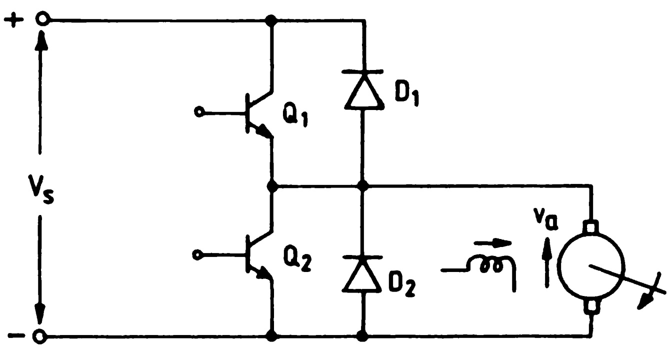


Fig. 3.17 Two-quadrant dc chopper-fed drive.

For four-quadrant operation, four transistors along with four diodes (Fig. 3.18) are needed. For quadrant I (forward motoring) operation, when the transistors Q_1 and Q_2 are turned on together, it is ON time as stated earlier. During OFF time, either Q_1 or Q_2, or both Q_1 and Q_2 can be turned off. If Q_1 is off, the motor current decreases through Q_2 and D_4. If both Q_1 and Q_2 are off, the current decreases through D_3 and D_4.

For quadrant II operation, i.e. forward (regenerative) braking, with the transistors Q_1, Q_2 and Q_3 being off, the transistor Q_4 is turned on. The current flows through Q_4 and D_2. When Q_4 is turned off, the current flows through D_1 and D_2, thus returning energy to the supply.

Similarly, quadrant III (reverse motoring) and quadrant IV (reverse braking) operations take place. For quadrant III, both the transistors Q_3 and Q_4 can first be turned on and then turned off, with the current passing through the diodes D_1 and D_2. Other forms of operation are also possible.

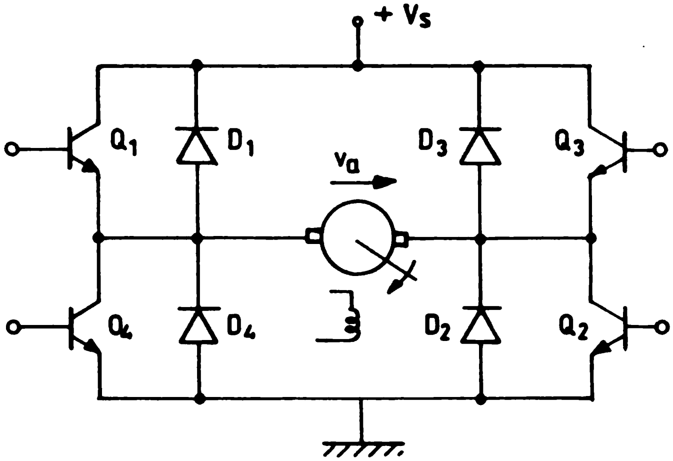


Fig. 3.18 Four-quadrant dc chopper-fed drive.

Only the transistor Q_2 can be turned on and then off for quadrant IV operation with other transistors Q_1, Q_3 and Q_4 remaining off. Both these cases are on the same lines as stated earlier.

3.4.3 Bridge Rectifier-fed Drive

One-quadrant operation

The half-controlled bridge rectifiers (converters), both single- and three-phase (Fig. 3.11, 3.13) can be used for one-quadrant (quadrant I) operation as the output voltage is positive and varies from 0 to $(V_{dc})_{max}$ when the delay angle α is varied from π to 0. The voltage V_{dc} is maximum at $\alpha = 0$, and zero (0) at $\alpha = \pi$. If used for armature voltage control of dc separately excited motors with the field current kept constant, the speed can be varied from zero to rated value in the forward direction. As the current is unidirectional, the developed electromagnetic torque is positive. Normally, the current is continuous, due to the presence of a freewheeling diode, even with values of inductance in the armature circuit lower than the value needed to make the current continuous in the full-controlled bridge rectifier. This is mostly applicable to the single-phase bridge rectifier, where the ripple is more than that in a three-phase bridge rectifier. The current can be discontinuous, especially at low speed with lower values of load torque. This has been described for single-phase full-controlled bridge rectifiers in the following subsection. The field can be supplied either from a constant voltage dc supply in series with a resistor or from a half-controlled single-phase bridge rectifier fed from ac supply, so as to vary the field current as needed.

Two-quadrant operation

The output voltage in the case of a full-controlled bridge rectifier (converter) varies from $-(V_{dc})_{max}$ to $+(V_{dc})_{max}$. At delay angle, $\alpha = 0$, the output voltage is maximum (positive), decreasing to zero at $\alpha = \pi/2$. If the delay angle is increased to π, the output voltage reaches its maximum (negative) value. So for $\pi/2 > \alpha > 0$, the output voltage is positive, and for $\pi > \alpha > \pi/2$, the output voltage is negative. This is valid when the current is continuous. The derivation of output voltage and current is shown. As the current is unidirectional (positive) for the full-controlled bridge rectifiers, whether it is single- or three-phase (Figs. 3.12 and 3.14), the operation will be either in quadrant I (forward motoring) or in quadrant IV (reverse braking). When the voltage is positive ($\pi/2 > \alpha > 0$), it is the motoring operation in the forward direction. When the voltage is negative ($\pi > \alpha > \pi/2$), it is the braking (regenerative) operation, but the motor must rotate in the reverse direction so that the output voltage is nearly equal to the emf developed in magnitude and sign, though the motor is now acting as a generator. This takes place in a hoist/crane in case of overhauling load, when the load is being lowered. Normally, if the motor is to be braked in the forward direction, quadrant II operation is needed, as the armature current is to be reversed. For this purpose as described later, reversing contactors either in the armature or in the field circuit, or two full-controlled bridge rectifiers in back-to-back connection for the armature circuit, which is called the dual converter, are used.

Discontinuous conduction

The armature of a dc separately excited motor is supplied from a single-phase full-controlled (thyristor) bridge rectifier (Fig. 3.12), while the field is fed from an uncontrolled (diode) bridge rectifier. The field current is kept constant. The equation of the voltage in the armature circuit is given by

$$R_a i_a + L_a \frac{di_a}{dt} + E = V_m \sin\theta \tag{3.10}$$

Depending on speed and load torque, especially low speeds combined with low values of load torque, the armature current is discontinuous. If delay angle is $\theta(\omega t) = \alpha$($\alpha$ will be large for low speeds and also as the input voltage balancing back emf developed would be low), the armature current is given by

$$i_a = \frac{V_m}{Z}\sin(\theta - \phi) - \frac{E}{R_a} + Ae^{-\frac{R_a\theta}{\omega L_a}} \tag{3.11}$$

The initial condition is

$$i_a = 0 \quad \text{at } \theta = \alpha$$

From this condition, using Eq. (3.11), A is obtained as

$$A = \left(-\frac{V_m}{Z}\sin(\alpha - \phi) + \frac{E}{R_a}\right)e^{\frac{R_a\alpha}{\omega L_a}} \tag{3.12}$$

Using Eq. (3.12) in (3.11), we get

$$i_a = \frac{V_m}{Z}\sin(\theta - \phi) - \frac{E}{R_a} + \left(-\frac{V_m}{Z}\sin(\alpha - \phi) + \frac{E}{R_a}\right)e^{-\frac{R_a(\theta - \alpha)}{\omega L_a}} \tag{3.13}$$

The current again goes to zero (0) at $\theta = \beta$, $\beta < (\pi + \alpha)$. The angle β is termed *extinction angle.* Substituting the condition in Eq. (3.13), we get

$$0 = \frac{V_m}{Z}\sin(\beta - \phi) - \frac{E}{R_a} + \left(-\frac{V_m}{Z}\sin(\alpha - \phi) + \frac{E}{R_a}\right)e^{\frac{R_a(\beta - \alpha)}{\omega L_a}} \tag{3.14}$$

Solving Eq. (3.14), β can be determined for given values of E and α. If the average value of i_a over the interval $(0, \pi)$ or (α, β) is obtained, the developed electromagnetic torque can be computed. The back emf (E) or the input dc voltage can also be computed. As the time interval is less than (π/ω) or $(T/2)$ $(T = 1/f)$, the motor speed is taken as constant with the mechanical time constant being large as compared to this time interval. The limiting case of discontinuous conduction is obtained, when for particular values of speed (back emf E) and delay angle α, the extinction angle β is given by $(\pi + \alpha)$. This is reached for particular values of speed and torque with the input (ac) supply voltage being kept constant at $V = V_m/\sqrt{2}$. If the delay angle α is decreased further, the speed increases as also the back emf E. The current remains continuous with increase in values of developed or load torque as the average current increases too. Thus, for continuous conduction, the speed must be higher with the developed or load torque also being higher.

Four-quadrant operation (Dual converter)

In reversible drives used for four-quadrant operation, the current is reversed either in the armature or in the field circuit. As the time constant in the armature circuit is lower, either a full-controlled bridge rectifier with a reversing contactor in the armature circuit (Fig. 3.19), or two full-controlled bridge rectifiers (converters) in back-to-back connection called the dual converter (Fig. 3.20), can be used to obtain four-quadrant operation of the drive. This scheme can be compared to the Ward-Leonard method (Section 3.2.1) suitable for four-quadrant operation. It is, therefore, also called the static Ward-Leonard scheme. The circuit shown in Fig. 3.20a is operated in noncirculating current mode, while for circulating current mode the circuit has to be slightly altered as shown in Fig. 3.20b.

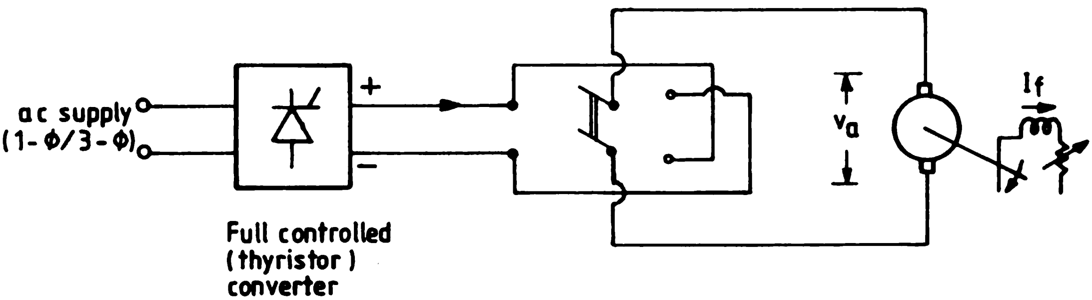
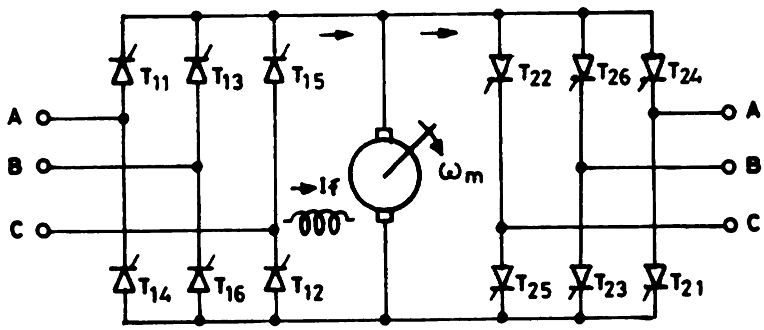
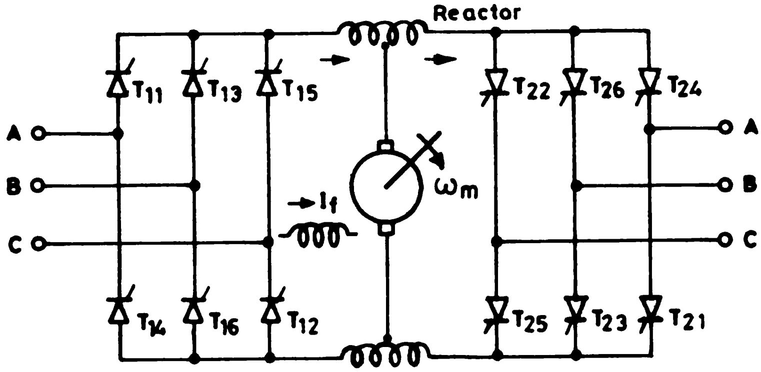


Fig. 3.19 One full-controlled converter with a reversing contactor in the armature for dc reversible drives (four-quadrant).

The bridge (#1) is used with $\pi/2 > \alpha > 0$ for quadrant I (forward motoring) operation. For quadrant II (forward braking) operation, the bridge (#2) is used with $\pi > \alpha > \pi/2$. The motor is run in the forward direction (quadrant I) at a certain speed with delay angle α_1. For braking operation in quadrant II, the delay angle must be set at $(\pi - \alpha_1)$ so as to have the same output voltage $(1.35\ V_L \cos \alpha_1)$ in the bridge (#2) to balance the induced emf in the motor as the speed remains the same. The motor is run in the reverse direction (quadrant III) using the bridge (#2) with $\pi/2 > \alpha > 0$. For braking the motor in the reverse direction (quadrant IV), the bridge (#1) is used with $\pi > \alpha > \pi/2$. Thus, four-quadrant operation is obtained. The same changeover can also be effected by using reversible contactors in the armature circuit with a full-controlled bridge rectifier as stated earlier, but the time delay is more due to contactor operation.



(a) Non-circulating current mode.



(b) Circulating current mode.

Fig. 3.20 Dual converter.

In the non-circulating current mode, only one bridge is operated at a time. So, when the motor is to be stopped (braked), the firing pulses to the thyristors in the bridge (#1) conducting at that time are withdrawn. Then, after waiting for one full cycle, the firing pulses are fed to the thyristors in the bridge (#2) to bring it to the conducting state for the braking operation. As two bridges must not conduct simultaneously, we must make sure that the thyristors in the outgoing bridge (#1) are off, before bringing the incoming bridge (#2) to conduction state. For this purpose, we have to wait for one full cycle, so that no short circuit takes place due to two bridges conducting at the same time. The only limitation in this mode is the time delay, needed to bring the incoming bridge to conducting state.

In the circulating current mode, both bridges conduct simultaneously, with the bridge (#1) taking the current through the motor in motoring/braking mode, while in the bridge (#2) the firing angle is so adjusted that only a small value of current, flows which can keep it in conduction. As the two voltages do not match from instant to instant, therefore inductors are needed to reduce the circulating current. The limitations in this mode are the losses that take place in the inductors and, the second conducting bridge. But the time delay is eliminated as both bridges conduct during the same time. The output voltages of both the bridges are nearly the same except for the voltage drop in the motor as shown earlier.

In all the above cases, either single-phase or three-phase bridge rectifiers can be used. The output voltage is given in Section 3.5.1. Normally, for dc motor drives of lower ratings, single-phase bridge rectifiers are used, while three-phase bridge rectifiers are preferred for drives with higher ratings. The control circuit used for triggering the thyristors is simpler for single-phase bridge rectifiers, while the circuit complexity, using inverse cosine firing control scheme, increases for three-phase rectifiers.

Example 3.5

The speed of a separately excited dc motor is controlled by a single-phase full-controlled bridge rectifier (converter), with the field also being controlled by a full converter. The field current is set to the maximum possible value. The supply voltage to the armature and field converters is 220 V, 50 Hz. The armature resistance R_a is 0.2 Ω, the field circuit resistance R_f is 150 Ω and the motor voltage constant, $k' = 1.1$ (V/A) · (rad/s) or (V · s)/(A · rad). The armature current I_a corresponding to the load demand is 25 A. The viscous friction and no load losses are negligible. The inductance of the armature and field circuits is sufficient to make the armature and field currents continuous and ripple free.

(i) If the delay angle of the armature converter is $\alpha_a = 45°$, and the armature current is $I_a = 25$ A, determine

(a) the torque developed by the motor,

(b) the speed, and

(c) the input power factor (pf) of the drive.

(ii) If the delay angle of the armature converter is changed to $\alpha_a = 0°$ for the same developed torque, determine

(d) the motor speed.

(iii) For the same load demand as in the earlier cases, determine

(e) the delay angle of the field converter, if the speed has to be increased to 1500 rpm.

Solution

$V = 220$ V, $I_a = 25$ A, $R_f = 150$ Ω, $R_a = 0.2$ Ω, $k' = 1.1$ (V · s)/(A · rad)

(i) For the first two cases, the maximum field voltage for delay angle $\alpha_f = 0°$ is

$$V_f = \frac{2\sqrt{2}\,V}{\pi} = \frac{2\sqrt{2} \times 220}{\pi} = 198 \text{ V}$$

The field current is $I_f = \dfrac{V_f}{R_f} = \dfrac{198}{150} = 1.32$ A

The motor voltage constant for this value of field current is

$$k = k'I_f = 1.1 \times 1.32 = 1.452 \text{ V} \cdot \text{s/rad}$$

(a) The torque developed by the motor

$$T_d = kI_a = 1.452 \times 25 = 36.3 \text{ N} \cdot \text{m}$$

The unit of k can be written as N · m/A or V · s/rad, as both are same.

(b) The armature voltage for delay angle $\alpha_a = 45°$

$$V_a = \frac{2\sqrt{2}\,V}{\pi}\cos\alpha_a = \frac{2\sqrt{2}\times 220}{\pi}\cdot\cos 45° = 198\times\cos 45° = 140\text{ V}$$

The back emf $E_{b1} = V_a - I_aR_a = 140.0 - (25.0\times 0.2) = 135.0$ V

$$\omega_1 = \frac{E_{b1}}{k} = \frac{135.0}{1.452} = 93.0\text{ rad/s},\qquad N_1 = \frac{60\omega_1}{2\pi} = \frac{60\times 93}{2\pi} = 888\text{ rpm}$$

(c) Assuming lossless converters, the total input power from the supply is

$$P_i = V_aI_a + V_fI_f = (140.0\times 25.0) + (198\times 1.32) = 3.76\text{ kW}$$

The input current to the armature converter for a highly inductive load is constant.

The rms value $I_{sa} = I_a = 25$ A

The rms value of the input current to the field converter $I_{sf} = I_f = 1.32$ A

The effective rms supply current is

$$I_s = \sqrt{(I_{sa})^2 + (I_{sf})^2} = \sqrt{(25)^2 + (1.32)^2} = 25.035\text{ A}$$

The input kVA rating is

$$VI = V_sI_s = 220\times 25.035\times 10^{-3} = 5.508\text{ kVA}$$

Neglecting the ripples, the input power factor (pf) is

$$\text{pf} = \frac{P_i}{VI} = \frac{3.76}{5.508} = 0.68\text{ (lag)}$$

(ii) (d) For delay angle, $\alpha_a = 0°$, $V_{a2} = 198$ V (as shown earlier) being also the same. Torque being the same, the current remains the same, the field current being also the same.

The back emf $E_{b2} = 198 - (25.0\times 0.2) = 193$ V

$$\omega_2 = \frac{E_{b2}}{k} = \frac{193}{1.452} = 133\text{ rad/s},\qquad N_2 = \frac{60\times 133}{2\pi} = 1270\text{ rpm}$$

(iii) (e) $N_3 = 1500$ rpm, $\quad \omega_3 = \dfrac{2\pi N_3}{60} = \dfrac{2\pi\times 1500}{60} = 157.1$ rad/s

$$E_{b2} = K'I_{f3}\omega_3 = 1.1\times 157.1\times I_{f3} = 193\text{ V}$$

$$I_{f3} = \frac{193}{1.1\times 157.1} = 1.117\text{ A}$$

$$V_{f3} = I_{f3}R_f = 1.117\times 150 = 167.6\text{ V}$$

Also

$$V_{f3} = 198 \cdot \cos \alpha_f = 167.6$$

or

$$\cos \alpha_f = \frac{167.6}{198} = 0.846$$

Therefore, the delay angle of the field converter $\alpha_f = 32.2°$

Example 3.6

If the polarity of the motor back emf in Example 3.5 case (i) with delay angle $\alpha_a = 45°$, is reversed by reversing the polarity of the field current, determine

(a) the delay angle of the armature converter α_a to maintain the armature current at the same value of $I_a = 25$ A, and

(b) the power fed back to the supply due to regenerative braking of the motor.

Solution

(a) From case (i) part (b) of Example 3.5, the back emf at the time of polarity reversal is

$$E_b = 135.0 \text{ V, and after polarity reversal, } E_b = -135.0 \text{ V}$$

$$V_a = E_b + I_a R_a = -135.0 + (25.0 \times 0.2) = -130.0 \text{ V}$$

$$V_a = 198 \cos \alpha_a = -130.0 \text{ V}$$

or

$$\cos \alpha_a = -\frac{130.0}{198} = -0.657$$

Therefore, the delay angle of the armature converter $\alpha_a = 131.0°$.

(b) The power fed back to the supply, $P_a = V_a I_a = 130.0 \times 25.0 = 3.25$ kW.

Example 3.7

A 200 V, 60 A dc series motor has armature and series field resistance of 0.06 Ω and 0.04 Ω, respectively. Running on no load with field winding connected to a separate dc source, the motor gave the following magnetization characteristic at 500 rpm:

Field current, A	10	20	30	40	50	60	70
Terminal voltage, V	53	98	125	142	153	162	168

The motor is controlled in regenerative braking by a chopper with input voltage of 200 V.

(a) Calculate the motor speed for a duty ratio of 0.4 at motor braking torque equal to rated value.

(b) Calculate the maximum allowable motor speed for maximum permissible values of current of 60 A and duty ratio of 0.9.

(c) What resistance must be inserted in the armature circuit for the motor to run at 800 rpm without exceeding armature current beyond 60A? The duty ratio of the chopper has a range from 0.1 to 0.9.

(d) Calculate the value of the diverter resistance to be connected in parallel with the series field to run the motor at 800 rpm without exceeding the armature current beyond 60 A.

Solution

(a) $\alpha_1 = 0.4$, $I_{a1} = 60$ A, $r_m = r_a + r_{se} = 0.06 + 0.04 = 0.1\ \Omega$, $V = 200$ V

The motor is braked by regenerative braking, i.e. runs as a generator.

$$E_1 = \alpha_1 V + I_{a1} r_m = (0.4 \times 200) + (60.0 \times 0.1) = 86 \text{ V}$$

For $I_{a1} = 60$ A and $N_2 = 500$ rpm, $E_2 = 162$ V (from the magnetization curve)

$$\text{Motor speed, } N_1 = \frac{E_1 N_2}{E_2} = \frac{86 \times 500}{162} = 265 \text{ rpm}$$

(b) $\alpha_{max}(\alpha_2) = 0.9$, $I_{a2} = 60$ A (current remains the same)

$$E_3 = (0.9 \times 200) + (60.0 \times 0.1) = 186 \text{ V}$$

$$\text{Motor speed, } N_3 = \frac{186 \times 500}{162} = 574 \text{ rpm}$$

(c) $N_4 = 800$ rpm, $\alpha_{max}(\alpha_3) = 0.9$, $I_{a3} = 60$ A

$$E_4 = \frac{N_4}{N_2} E_2 = \frac{800 \times 162}{500} = 259.2 \text{ V}$$

$$R_{rh} + 0.1 = \frac{E_3 - \alpha_3 V}{I_{a3}} = \frac{259.2 - (0.9 \times 200)}{60} = 1.32\ \Omega$$

or

$$R_{rh} = 1.32 - 0.1 = 1.22\ \Omega$$

(d) Firstly, it is assumed that after using the diverter, the combined resistance remains the same at 0.1 Ω.

$$\alpha_{max}(\alpha_4) = 0.9,\ I_{a4} = 60 \text{ A}$$

$$E_5 = 186 \text{ V (same as } E_3)$$

This is the back or generated emf developed by the machine at 800 rpm.

$$\text{At } N_2 = 500 \text{ rpm, } E_6 = \frac{500 \times 186}{800} = 116.25 \text{ V}$$

The current required in the series field

$$I_{se6} = \frac{E_6}{E_2} I_{a1} = \frac{116.25 \times 60.0}{162.0} = 43 \text{ A} \quad (\because I_{se1} = I_{a1})$$

The current through the diverter = $I_{a6} - I_{se6} = 60 - 43 = 17$ A ($\because I_{a6} = I_{a1}$)

$$17.0 \times R_d = 43.0 \times 0.04 = 1.72 \text{ V}$$

or

$$R_d = \frac{1.72}{17.0} \cong 0.1 \ \Omega$$

The equivalent resistance of the total circuit changes from 0.1 Ω to 0.089 Ω. This value has not changed much as $R_d \cong 0.1\ \Omega$.

3.5 DC MOTOR CONTROLLERS

3.5.1 Controllers for DC Shunt Motors

The shunt field current is fed from a diode bridge rectifier, in series with a resistance. The armature voltage is varied by controlling the firing angle of the thyristor bridge rectifier (Fig. 3.21). The speed variation is sensed by an ac tachogenerator. The voltage signal, thus, received is rectified and after proper smoothening, is compared with the reference voltage. The error voltage is then fed to the base of the first control transistor, which is called the speed amplifier. During starting, the IR compensation circuit is kept inactive. The speed control is effected through the inner (current) and outer (speed) regulating loops. The tachogenerator feedback is included in the outer loop and current feedback in the inner loop. The regulator may also be provided with soft start and current limit.

3.5.2 Microprocessor-based Controller for DC Motor Drives

Single-phase thyristor bridge converter-fed dc motor drive

A controller (Fig. 3.22) for dc motor drive using an 8-bit microprocessor (8085) is described. A thyristor bridge converter is used to feed the armature circuit of the dc shunt motor, while the field is supplied from a diode bridge. An open loop controller is described first. A programmable timer/counter chip (8253) is also used in the processor-based system for the control of time delay as needed for armature voltage control with the field current being constant. The different blocks needed for this purpose are: (a) the synchronizing module, and (b) the driver module.

For the output of the thyristor bridge converter to be varied in order to control the motor speed, the thyristors need to be turned on after certain time delay, when the zero crossing of the input voltage is detected. The field current is kept constant. The synchronizing module (Fig 3.23a) consisting of an operational amplifier, used as comparator, detects both the positive and negative zero crossing points. Two timer/counters (Fig. 3.23b) of the chip (8253) are used in monostable mode (mode 1). In this mode, the gate input is used as the synchronizing or trigger signal, when the time delay starts. The gate is connected to the output of the synchronizing module. The output of the timer goes low when the gate input goes high. Before this the count register must be loaded with a proper value. The register starts downcounting from the time of synchronization. The output goes high after the count register decrements to zero. Thus, the specified time delay is reached (Fig. 3.23c). The clock of the timer/counter chip is quite high being of the order of 1.5 MHz,

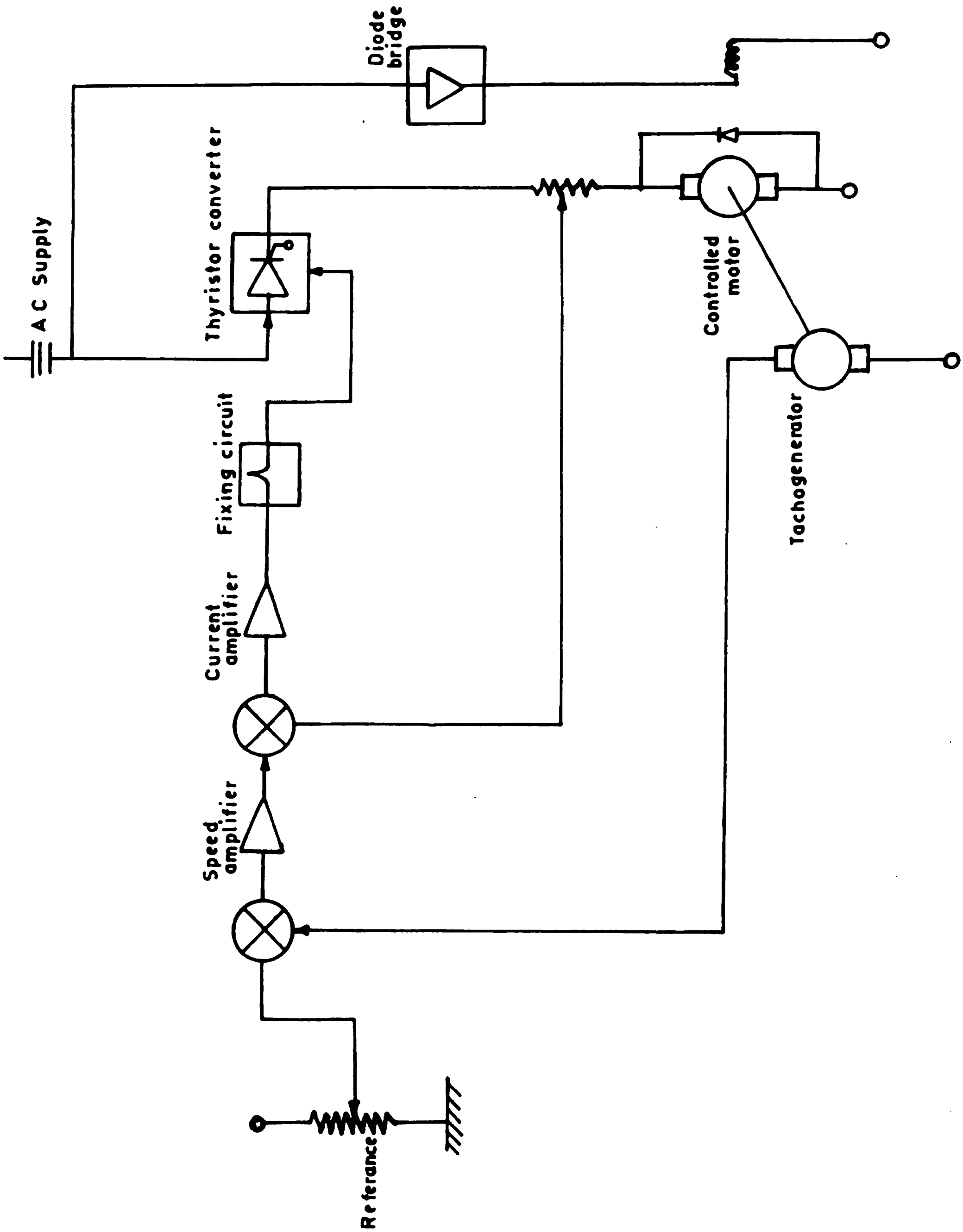


Fig. 3.21(a) Controller for dc shunt motors—Schematic diagram.

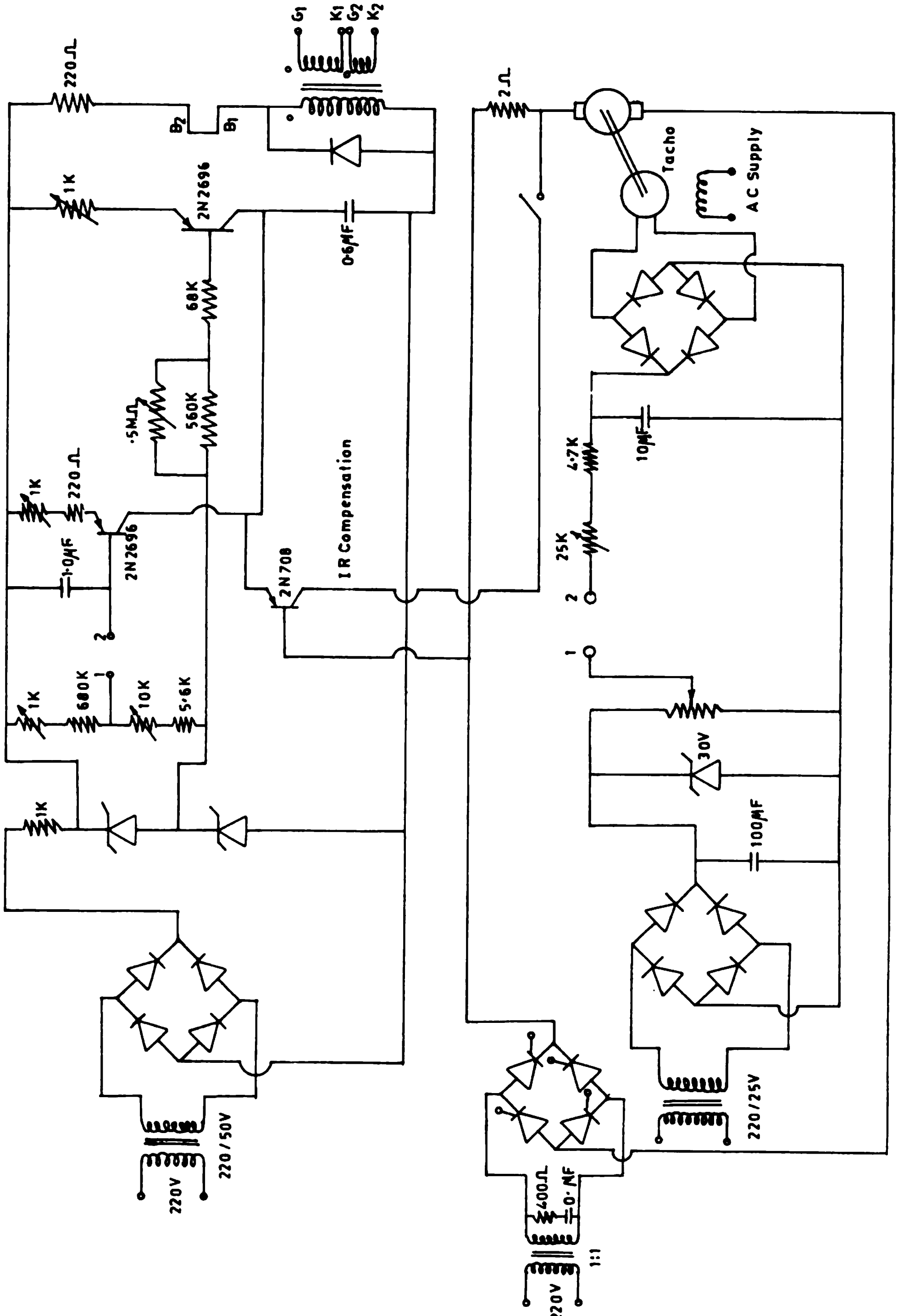


Fig. 3.21(b) Controller for dc shunt motors—Single-phase converter control.

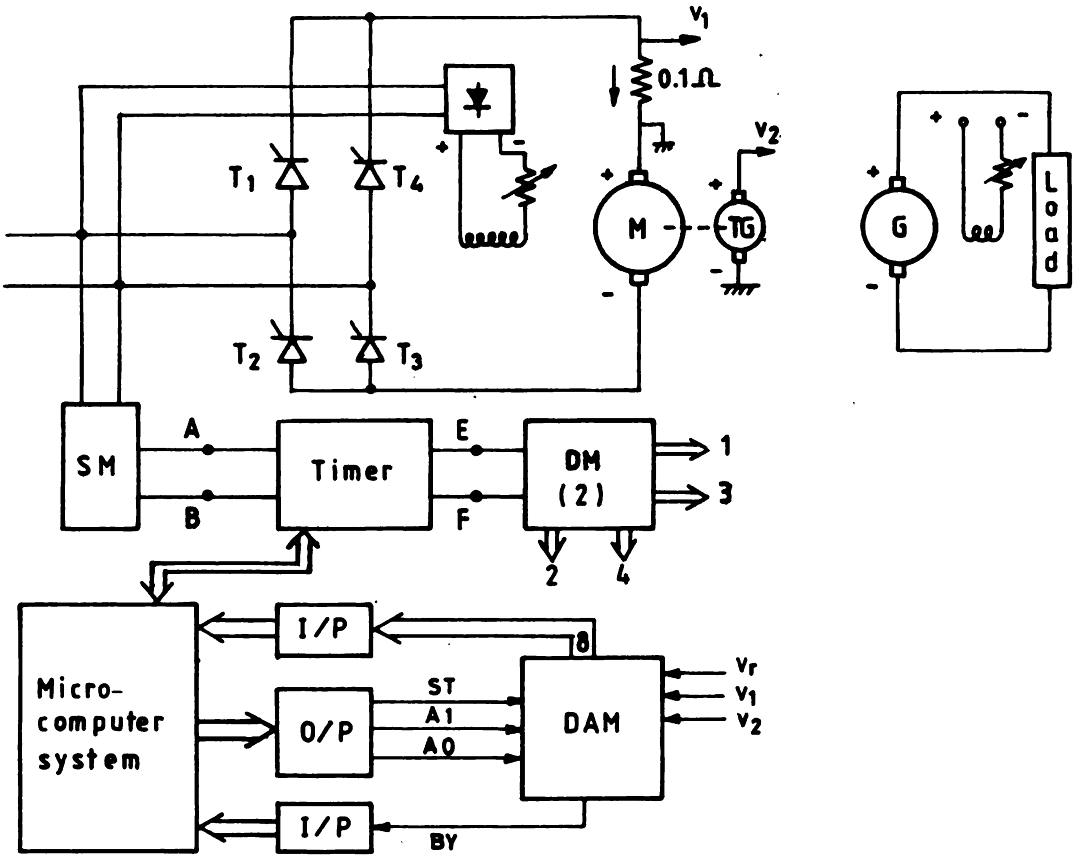


Fig. 3.22 Microprocessor controlled thyristor bridge dc motor drive.

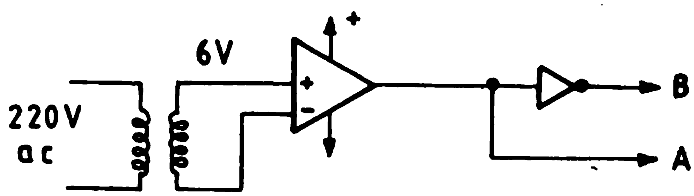


Fig. 3.23(a) Synchronizing module.

obtained from the 3 MHz clock signal of the 8085 processor so as to get a high resolution of the time delay. For example, to create a delay of 45°($\pi/4$), the delay needed is only 2.5 ms, assuming supply frequency of 50 Hz. The time period, T [360° (2π)] is 20 ms. For this delay, the count register needs to be loaded with 3750_{10}. Two timer/counters are needed, one for the positive zero crossing signal, and the other for the negative zero crossing signal.

The driver module (Fig. 3.23d) consists of one/two transistors used as amplifier, gates as needed along with pulse transformers, the output of which is fed to the gates of the thyristor pair. This is needed, as the timer output cannot be directly used for connection to the gates of the thyristors.

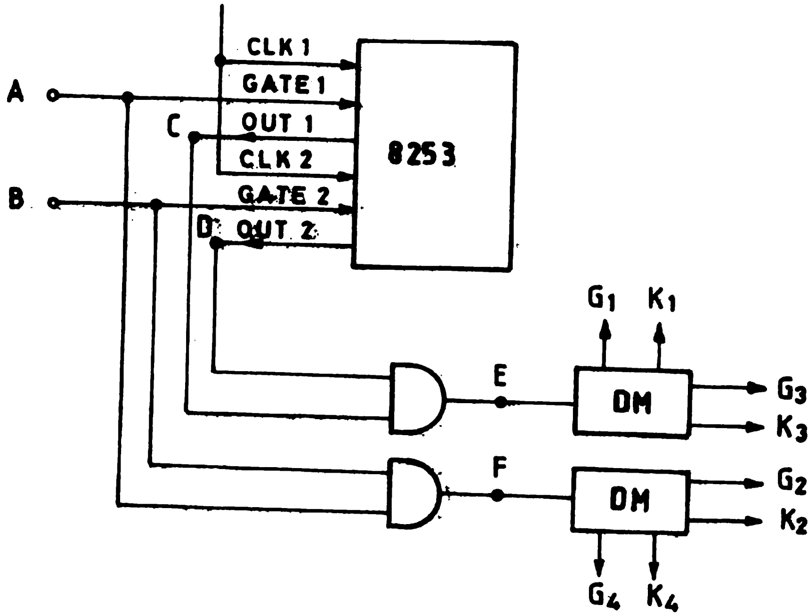


Fig. 3.23(b) Circuit connections for timer/counter chip (8253).

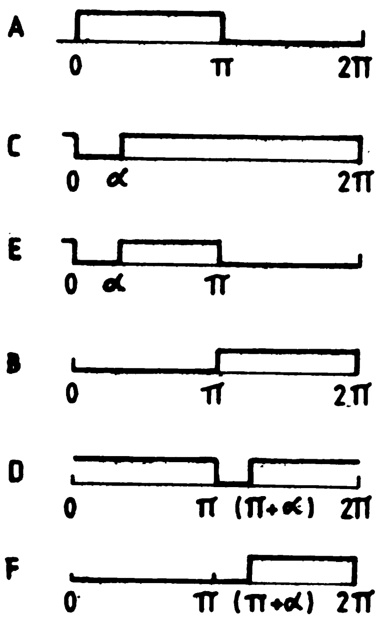


Fig. 3.23(c) Waveforms for Figs. 3.23(a) and (b).

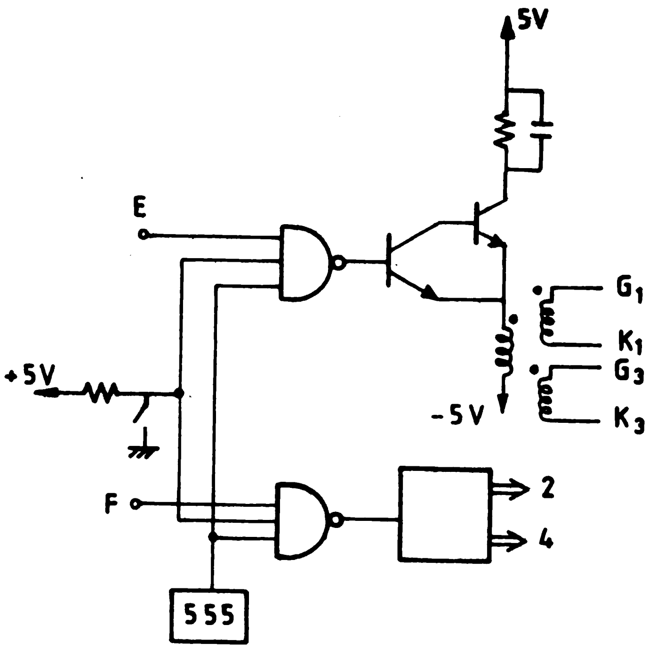


Fig. 3.23(d) Driver module.

If the counter (#1) is used, the programs required are:

(1) initialization part, and (2) time delay generation using counter (#1) of 8253 counter/timer chip in monostable mode (mode 1). The port numbers are 11H for count register of the counter (#1) and 13H for mode register. To initialize the counter (#1) in mode 1, the program is

```
LPX:  MVI   A, 72H
      OUT   13H      ;72H [0111 0010B] is loaded in mode register (counter #1 in
                     ;mode 1, binary)
```

For time delay generation, the count register (16-bit) is loaded with a 16-bit data, which is loaded in sequence, MSB first, followed by LSB via the same port no., only one is used for this case in the counter chip. The program for this case is

```
LPY:  MVI   A, XXH
      OUT   11H      ;XXH is loaded in MSB (8-bit) of count register (#1)
      MVI   A, YYH
      OUT   11H      ;YYH is loaded in LSB (8-bit) of count register (#1)
```

So, XXYYH is loaded in count register of counter (#1). As auto-reload feature is used, the value is loaded only once after initialization. But, as the data is changed in this case after computation in each step as shown later in Section 3.6, the value (XXYYH) is modified at a certain interval, say, after each sampling as needed.

Only the program for counter (#1) is described. Other counter (#2) is also used, with the program being identical, except that the port no. (12H) for count register of counter (#2) and the initialization part, i.e. the data (B2H) to be loaded in mode register (13H), are changed. This part has been described in the subsection for dc thyristor chopper.

The scheme can be modified by adding an extra circuit such that only one counter/timer of the programmable counter chip (8253) is used in the manner described in order to produce time delay after each zero crossing point, both in the positive and negative half.

The time delay can also be produced by software by using the synchronizing and driver modules, as described earlier. After zero crossing in any half cycle has been detected, the 'DELAY' subroutine is used to produce the required time delay and at the end of which the corresponding thyristor pair is triggered or turned on. The thyristor pair is turned off or pulses at their gates are stopped, when the next zero crossing is detected. The scheme uses minimum hardware, needing no programmable counter/timer chip, but the main drawback is the use of CPU time to produce time delay, whereas earlier the CPU was free to perform any other tasks such as control law computation or fault diagnostics in the system. The first program described is used to detect the two zero crossing points. The two signals from the synchronizing module, both positive and negative zero crossing signals, are fed to bits 0 and 7 of Port B, which is used as the input port of the programmable I/O port chip (8255) as described in the section on Data acquisition model (DAM). The program is

```
BCK1: IN    02H
      RAR
      JNC   BCK1     ;positive zero crossing point is detected, as bit 0 at Port B goes
                     ;from low (0) to high (1)
```

The bits 0 and 1 of Port C, configured as output port of the same chip (8255), are connected to the inputs of the two driver modules needed to turn on the two thyristor pairs. The odd pair of thyristors is turned on after positive zero crossing point, and even pair after negative zero crossing point. It is to be noted that the two pairs of thyristors must be turned off, or pulses at the gates of thyristors must be stopped, as soon as any of the two zero crossing points is detected. The complete program is given below:

```
LPA:   MVI   A, 01H
       PUSH  PSW       ;Move 01H (0000 0001B) into stack
LPB:   POP   PSW
       SUI   01H
       JZ    LPC       ;If result is zero, check for negative zero crossing point
       MOV   B, A      ;Move 01H into reg. B, and later in the program into stack (this
                       ;takes place only after the loop is completed)
BCK1:  IN    02H
       RAR
       JNC   BCK1      ;Positive zero crossing point is detected
       JMP   LPD
LPC:   MVI   B, 02H    ;Move 02H (0000 0010B) into reg. B, and later into stack
BCK2:  IN    02H
       RAL
       JNC   BCK2      ;Negative zero crossing point is detected
LPD:   XRA   A
       OUT   03H       ;Move 00H (0000 0000B) to Port C, such that pulses to the
                       ;gates of the two thyristor pairs are stopped to turn them off
       PUSH  B         ;Move 01H or 02H into stack as described earlier
LXI    D, ZZWWH        ;Load register pair (D, E) with suitable 16-bit data
                       ;(ZZWWH) to produce the required time delay in open loop
       CALL  DELAY
       POP   PSW
       OUT   03H       ;Depending on whether positive or negative zero crossing
                       ;point is detected, the corresponding thyristor pair is triggered
       PUSH  PSW
       JMP   LPB
```

A standard time delay subroutine using a 16-bit register pair (D, E) is used. As the decrement register pair (DCX D) instruction does not affect zero flag, two instructions, MOV A, D and ORA E are used, in which the data (bits) stored in D and E registers are ORed to check, if the data in both the registers or register pair (D, E) has decreased to zero (0000H). When the content of the register pair (D, E) is zero, the program comes out of the time delay loop, and the preset time delay is reached at that time. The program for DELAY subroutine is

```
DELAY: DCX   D
       MOV   A, D
       ORA   E
       JNZ   DELAY    ;If the contents of the register pair (D, E) is not zero, it is checked
                      ;again, till it reaches zero, when the loop is completed
       RET
```

The program as described above can be modified, so as to reduce the memory needed and also the execution time. It can also be simplified, or increased in size requiring more time for execution. Care should be taken to see that the execution time needed for this part of the program is taken into account properly, while calculating the data to be stored in the register pair for the required time delay. Otherwise, the time delay would be more than what is required or preset. This is the case for on-line or real-time programs.

This is not so in the case of the previous program using a counter chip, as the program is very short and hence the execution time is low. But for very precise time delay generation, the execution time can be taken care of while computing the data to be loaded in the count registers of the counters. This is also valid for the program in the case of a dc thyristor chopper using programmable counters as described later.

Data acquisition module (DAM)

For closed loop control, the states of the drive system such as motor speed, set (reference) speed and motor current are measured by a processor-based system. In this case, a dc tachogenerator is used to obtain a voltage proportional to speed. This voltage after filtering is fed to the input (#1) of an 8-channel analog multiplexer. The resistance (standard) in the armature circuit is used to measure the motor current. The voltage drop in the resistance after proper filtering is also fed to input (#2) of the multiplexer. The voltage proportional to set speed is scaled down to the range of dc tachogenerator output voltage and then fed to input (#0) of the multiplexer.

The data acquisition module (DAM) shown in Fig. 3.23e consists of an 8-channel analog multiplexer (MUX), a sample and hold (S/H) and an 8-bit analog to digital converter (ADC). The output of ADC is fed to an input port. The output lines used for control are the two channel selection

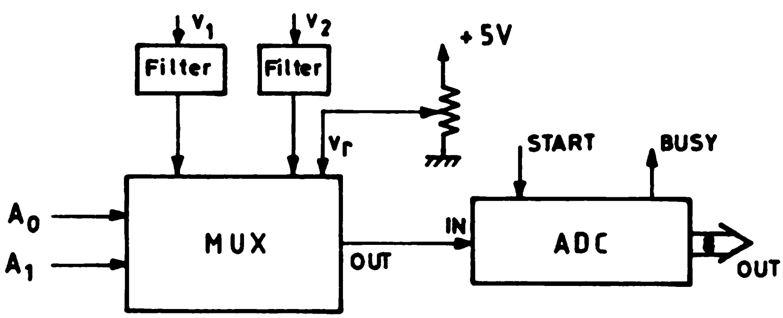


Fig. 3.23(e) Data acquisition module with filters.

lines of the multiplexer for four channels, one hold line for sample and hold and a start line for ADC. The input line is the busy line of the ADC used in flag-testing mode. The ADC used is of successive approximation type and is quite fast. To measure any input signal, the particular channel selection lines are selected such as 00B for channel #0 and the hold and start lines are made high. The busy line goes high immediately and goes low after some time, when the ADC has completed conversion of the input signal. This line is tested in the flag-testing mode and when the busy line is low, the output of the ADC is read by the processor. The input signals are read sequentially, three in this case. Only the channel selection lines are changed in each case and the outputs are stored in proper locations in the memory.

The Programmable Peripheral Interface (PPI) chip (8255), also termed Programmable Input/ Output (I/O) Port, is used for the purpose. The port no. of the control (mode select) register is 00H and Ports A, B and C are 01H, 02H and 03H, respectively. The output lines (8) of the 8-bit ADC are fed to Port A used as an Input Port (I/P). Port C is used as an Output Port (O/P), bits 7 and 6 for two channel selection lines of the MUX and bits 5 and 4 for sample and hold (S/H) and start line of ADC, respectively. Port B is also used as Input Port, bit 1 being connected to the busy line of the ADC. To select the three Ports, A, B and C as stated, the control register is loaded with the data 92H. Mode 0 is used for this purpose. The initialization part of the program is

```
LPT:  MVI   A, 92H
      OUT   00H      ;92H [1001 0010B] is loaded in the control register to select
                     ;Ports A and B as I/P and Port C as O/P of PPI (8255) chip
```

To measure any analog input data, say, channel #0, i.e. the set (reference) speed, the data to be outputted at Port C are 30H (0011 0000B), both S/H and start lines are made high (1), earlier these lines must be held to be low (0). The two channel lines are 00B. Earlier data at Port C is 00H (0000 0000B). Then, the busy lines are checked at bit 1 of Port B to find when it goes low in flag testing mode. Lastly, the data at Port A is read and stored in a memory location. The program is

```
LPS:  XRA   A
      OUT   03H      ;00H is outputted at Port C
      MVI   A, 30H
      OUT   03H      ;30H is outputted at Port C
BCK:  IN    02H
      ANI   02H
      JNZ   BCK      ;bit 1 at Port B is checked to find, when it is low (0); if it is
                     ;high (1), it is checked again
      IN    01H
      SHLD  ZZWWH    ;data at port A is stored in memory location (ZZWWH)
```

For channel #1, the initial and final data to be outputted are 40H (0100 0000B) and 70H (0111 0000B), respectively. The other parts remain the same except for the memory location, which has to be modified. Similarly, for other channels, say #2, the data need to be changed accordingly.

If this program is to be used along with the program for time delay generation and driver modules, which is connected to Port C (O/P) of I/O port chip (8255), then both the programs must be properly modified, as some other bits of Port C are used for this module (DAM) in the program.

Only one form of modification is described here. It may be stated here that some other program can also be used in this case. Any book or manual for Intel 8085 (8-bit) processor may be consulted.

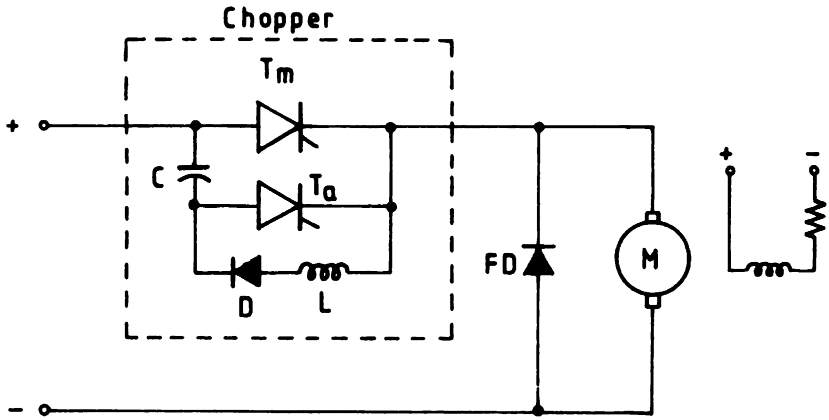
The data outputted at Port C must be stored at the memory location labeled PORTC (UUVVH) in the previous program starting from LPS: XRA A. The new program should replace the *first four* lines of the previous one. The instruction ANI 03H is used to preserve the bits 0 and 1 of output Port C. Then the data is initially outputted at Port C so as to specify the channel no. before the start signal is applied to ADC, and later the data is again outputted at Port C after adding 30H, as explained in the previous case. This is the modification needed for channel #0 as shown in the program below. For channel #1, the instruction ADI 40H must be inserted after ANI 03H, as the channel no. needs to be specified before the start signal is applied to ADC as stated earlier.

```
LPR:  LDA   PORTC
      ANI   03H      ;bits 1 and 2 of Port C are preserved, so as not to affect the other
                     ;program
      OUT   03H
      ADI   30H
      OUT   03H
      STA   PORTC
```

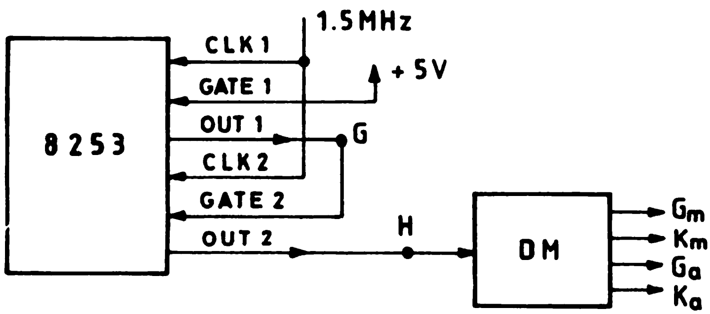
Thyristor chopper-fed dc motor drive

The thyristor chopper circuit (Fig. 3.24a) described in Section 3.4.2 produces a variable dc output voltage. This voltage as in the earlier case of the thyristor bridge converter can be used to control the speed of a dc shunt motor. Two timer/counters (Fig. 3.24b) of the chip (8253) are used to obtain the triggering signals as needed for the main and auxiliary thyristors of the chopper. One is used to generate a fixed frequency signal, either a square wave of period T or a single pulse at an interval of time period T. This timer/counter (#1) is operated in mode 3 or 2. In mode 3, a square wave of time period T is produced (Fig. 3.24c). In the second case (mode 2), a single pulse going from high to low for one clock period at an interval of time period T is obtained. The clock frequency of both timer/counters as in the earlier case is 1.5 MHz. If the chopping frequency is assumed as 200 Hz, the time period is 5 ms. The count register of timer/counter (#1) is loaded with 7500_{10} to produce the waveform with time period (T = 5 ms). Also, the gate is made high.

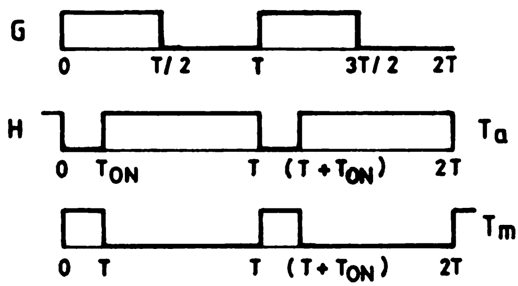
The other timer/counter (#2) is operated in mode 1 (monostable mode) as was the case earlier. The gate of this timer is connected to the output of timer (#1). At the time the output of timer (#1) goes from low to high, the output of timer (#2) goes from high to low. The count register must be loaded earlier. The register starts downcounting and when it reaches zero, the output changes from low to high and remains high till the next transition (low to high) occurs at the gate of timer (#2), i.e. output of timer (#1). This waveform (Fig. 3.24c) is used for firing the auxiliary thyristor and the inverted one for the main thyristor of the chopper. The time, during which the output of timer (#2) remains low, is T_{ON} (ON time) of the main thyristor. The main thyristor is triggered at time, $t = 0$, and the auxiliary one at $t = T_{ON}$. The cycle repeats itself. For proper operation of the chopper, the auxiliary thyristor is initially switched on to charge the commutating capacitor and after that the cycle begins. As in the earlier case, two driver modules (Fig. 3.24b) are needed so as to connect the timer outputs to the proper thyristor gates. If the duty cycle (α) varies from 0.1 (10%) to 0.9 (90%), the count value to be loaded varies from 750_{10} to 6750_{10}.



(a) Power circuit.



(b) Circuit connections for timer/counter chip (8253).



(c) Waveforms for Fig. 3.24b.

Fig. 3.24 Chopper controlled dc motor using counter/timer chip.

This scheme can also be used for a transistor chopper with high chopping frequency. Only the driver modules need to be modified so as to be able to drive the base of the power transistors to switch them ON and OFF as required. The chopping frequency can be altered using software only, changing the values to be loaded into the count registers of the timer/counters.

As stated earlier, the initialization part is described first. The counter (#1) is initialized in astable mode (mode 3) and counter (#2) in monostable mode (mode 1). Using the same port no. for the counter chip, the program is

```
LPU:  MVI  A, 76H
      OUT  13H     ;76H [0111 0110B] is loaded in mode register (counter #1 in
                   ;mode 3, binary)
      MVI  A, B2H
      OUT  13H     ;B2H [1011 0010B] is loaded in mode register (counter #2 in
                   ;mode 1, binary)
```

For loading the count registers (#1 and #2), the program starting from LPY: MVI A, XXH as stated in Section 3.5.2 is modified as shown below. The count register (11H) of counter (#1) is loaded with the value, 7500_{10} (1D 4CH) and the count register (12H) of the counter (#2) with the data as computed (shown in Section 3.6).

```
LPW:  MVI  A, 1DH
      OUT  11H
      MVI  A, 4CH
      OUT  11H     ;The data, 1D4CH is loaded in count register of counter (#1)
LPV:  MVI  A, XXH
      OUT  12H
      MVI  A, YYH
      OUT  12H     ;The data XXYYH is loaded in count register of counter (#2)
```

Closed loop control

For closed loop control, the control law as needed is implemented via software. A proportional-plus-integral (P-I) controller for speed control needs only two inputs—set speed and motor speed—measured by the module as described. The software developed computes the control input and then converts it to the angle or count value to be loaded in the count registers of the timer/counter chip. In most cases, especially for large step changes in set speed, the armature current is very large for dc shunt motors. So a current controller of P-type or P-I type is also implemented as an inner loop, while the speed controller is in the outer loop. The function of the current controller is to control the current to the safe limit and for this purpose, a limiter is placed at the output of the speed controller so that the current limit is never exceeded. Normally, the inner current loop acts much faster, while the outer speed loop is slower as compared to the inner loop.

3.6 STATE FEEDBACK CONTROLLER FOR DC MOTOR DRIVES

The state feedback controller for a dc motor speed control system is described in this section in brief. Before taking up an example, the theory is explained first.

The speed control system is modelled in state space form as

$$\begin{aligned} \dot{\underline{x}} &= A\underline{x} + B\underline{u} + E\underline{d} \\ \underline{y} - \underline{y}_r &= C\underline{x} - \underline{y}_r \end{aligned} \tag{3.15}$$

where

$\underline{x}$, $\underline{u}$, $\underline{d}$, $\underline{y}$, and $\underline{y}_r$ are the state, control (input), disturbance (input), output and reference input vectors of length n, m, q, p and p, respectively, and

A, B, C, and E are matrices of dimension $(n \times n)$, $(n \times m)$, $(p \times n)$, and $(n \times q)$, respectively.

In the extreme case, $\underline{d}$ and $\underline{y}_r$ have dynamics of their own and may be modelled as systems of exogenous origin. However, for most industrial problems, $\underline{d}$ is a constant vector (step type disturbances) and $\underline{y}_r$ is also constant giving rise to a set point (reference) regulation problem.

Mathematically speaking, the regulation problem with internal stability (RPIS) requires that

$\dot{\underline{x}} \to 0$ (stability of the states), and

$$\tilde{\underline{y}} = \underline{y} - \underline{y}_r \to 0 \text{ or } \underline{y} \to \underline{y}_r \text{ (output regulation) as } t \to \infty \tag{3.16}$$

Assuming $\underline{d}$ and $\underline{y}_r$ as constants, differentiation of Eq. (3.15) yields

$$\dot{\underline{z}} = \hat{A}\underline{z} + \hat{B}\underline{v} \tag{3.17}$$

where

$$\underline{z} \overset{\Delta}{=} \begin{bmatrix} \dot{\underline{x}} \\ \underline{y} - \underline{y}_r \end{bmatrix}, \quad \underline{v} \overset{\Delta}{=} \dot{\underline{u}}, \quad \hat{A} = \begin{bmatrix} A & 0 \\ C & 0 \end{bmatrix} \text{ and } \hat{B} = \begin{bmatrix} B \\ 0 \end{bmatrix} \tag{3.18}$$

If the eigenvalues of augmented system matrix $\hat{A}$ are made to lie in the open left half plane by the state feedback method or any other method, the RPIS is solved.

State feedback

The fundamental theorem of feedback control states that for a system

$$\dot{\underline{x}} = A\underline{x} + B\underline{u} \tag{3.19}$$

with a state feedback

$$\underline{u} = K\underline{x} \tag{3.20}$$

the closed loop system

$$\dot{\underline{x}} = (A + BK)\underline{x} \tag{3.21}$$

can have any pre-assigned set of eigenvalues, if and only if the pair (A, B) is controllable. The theorem gives rise to the famous pole assignment technique of linear system theory.

The formulae for pole assignment are:

Real pole (shifting from λ_i to λ_i')

$$k_i = \frac{(\lambda_i' - \lambda_i)}{(\underline{v}_i^T \underline{b}_j)} \tag{3.22}$$

where

k_i is the scalar gain,

$\underline{v}_i^T$ is the left eigenvector of A^T (right eigenvector of A) corresponding to λ_i, and

$\underline{b}_j$ corresponds to the control input used for shifting the eigenvalue, $j \le m$.

The poles are shifted one-by-one, and after each shifting the resulting closed loop system is taken as the base for next shifting.

The vector gain $\underline{k}_j^T$ is given by

$$\underline{k}_j^T = k_i \underline{v}_i^T \tag{3.23}$$

and develops into a gain matrix K, when multiple inputs are used for pole shifting.

Pair of complex poles (λ_i, λ_i^* shifted to λ_i', $\lambda_i'^*$)

$$k_i = \frac{(\lambda_i' - \lambda_i)(\lambda_i'^* - \lambda_i)}{(\underline{v}_i^T \underline{b}_j)(\lambda_i^* - \lambda_i)} \tag{3.24}$$

The resulting state feedback control is, however, real and given by

$$u_j = [k_i \underline{v}_i^T + k_i^* \underline{v}_i^{*T}]\underline{x} = \underline{k}_j^T \underline{x} \tag{3.25}$$

The amount of pole shift or placement of new poles is guided by engineering judgement and also, the input to be used for pole shifting must be properly chosen. The feedback gains should not be too high as the resulting control inputs would be high for physical systems.

For RPIS, the controllability conditions of the pair $(\hat{A}, \hat{B})$ may be broken up into the following simpler conditions:

(i) pair (A, B) is controllable, and

$$\text{(ii) rank} \begin{bmatrix} A & B \\ C & 0 \end{bmatrix} = (n + p) \tag{3.26}$$

A direct corollary of (ii) in Eq. (3.26) is $m \le p$, that is, the number of inputs m should be greater than or at least equal to the number of outputs p to be regulated.

For asymptotic regulation, a weaker condition of stabilizability is sufficient. Stabilizability requires that either

(i) the pair (A, B) is controllable, or

(ii) the uncontrollable modes are stable.

In this case, the control is not achieved in a finite time T, but at time, $T = \infty$.
The structure of the pole placement controller comes out as

$$\underline{u} = K_1 \underline{x} + K_2 \int (\underline{y} - \underline{y}_r) dt \tag{3.27}$$

where K_1 and K_2 are obtained by partitioning the K matrix of the control law of the augmented system (3.17)

$$\underline{v} = K\underline{z} \tag{3.28}$$

It is the combined state feedback and integral-of-output-error (IOE) control.

3.6.1 Combined Observation and Control

The problem of incomplete state feedback, or alternatively output feedback, is taken care of by combined observation and control. Provided the pair (C, A) is observable, the techniques of controller design by pole assignment can be extended to the design of Luenberger type full order observer. However, if some of the states are available directly as outputs or combination of outputs, a reduced order observer serves the purpose.

A reduced order observer is a dynamical system given by

$$\dot{\hat{\underline{\xi}}} = D\hat{\underline{\xi}} + F\underline{y}' + G\underline{u}' \tag{3.29a}$$

$$\underline{\xi} = L\underline{x} \tag{3.29b}$$

The L is a transformation matrix such that $[C' \ L]^T$ is invertible, where C' consists of linearly independent rows of C, and y changes correspondingly to y'. Thus, the order of the observer is given by $(n - r)$, where r is the rank of C, $r \leq p$. For the purpose of observation, no difference is made between control and disturbance inputs; B is extended to B' by inclusion of E.
Differentiating Eq. (3.29b) and using (3.19), we get

$$\dot{\underline{\xi}} = L(A\underline{x} + B'\underline{u}') \tag{3.30}$$

Subtracting Eq. (3.30) from (3.29a), we have

$$\dot{\hat{\underline{\xi}}} - \dot{\underline{\xi}} = D(\hat{\underline{\xi}} - \underline{\xi}) + (G - LB)\underline{u} + (FC' - LA + DL)\underline{x} \tag{3.31}$$

The problem of observability is stated as

$$\dot{\tilde{\underline{\xi}}} = \dot{\hat{\underline{\xi}}} - \dot{\underline{\xi}} \rightarrow 0, \text{ as } t \rightarrow \infty$$

For the error in estimate $(\hat{\underline{\xi}} - \underline{\xi})$ to decay

(a) $LA - DL - FC' = 0$ (3.32)

(b) the eigenvalues of D must lie in the open left half plane with the restriction $\{\lambda_i\}_D \neq \{\lambda_i\}_A$, otherwise F becomes trivial.

(c) $G = LB'$; however, not with any arbitrary choice of L. (3.33)

A choice of L is

$$L = \sum_{i=1}^{n} \underline{l}_i \, \underline{v}_i^T \tag{3.34}$$

where, $\underline{v}_i^T$ is as in (3.22).

The eigenvalues of D should be deep in the left half plane for a 'fast' observer. As in the controller design, a weaker condition of detectability is applicable here, if finite time observation is not required.

The $\underline{l}_i$'s of Eq. (3.34) are given by

$$\underline{l}_i = [\lambda_i I - D]^{-1} F[C'\underline{w}_i] \tag{3.35}$$

where, $\underline{w}_i$ is the left eigenvector of A corresponding to λ_i.

F is by choice.

The structure of the observer comes out to be the same as that of Kalman Filter, except that the gain is sub-optimal.

Combined observation and control may give rise to structural instability, which cannot be foretold. However, by trial and error this difficulty can be circumvented.

3.6.2 Feedforward Control

Dynamics of the system is controlled by pole assignment. However, for improvement in the transient performance, zeroes also need to be assigned. The age-old method of feedforward control, used by the process control engineers, does implicit zero assignment. Feedforward control from disturbance and reference inputs is very effective, particularly for complete disturbance rejection and quick following of command (reference input).

Assuming $\underline{d}$ and $\underline{y}_r$ as vectors of constants for ease of illustrating the technique and that the system has reached steady state, Eq (3.15) changes to

$$0 = A\underline{\bar{x}} + B\underline{\bar{u}} + E\underline{d} \tag{3.36}$$

$$0 = C\underline{\bar{x}} - \underline{y}_r$$

where $\underline{\bar{x}}$ and $\underline{\bar{u}}$ are the steady state values.

By condition (ii) of Eq. (3.26), this equation has a solution, which is given by

$$\begin{bmatrix} \underline{\bar{x}} \\ \underline{\bar{u}} \end{bmatrix} = -\left[G_f^T G_f\right]^{-1} \left[G_f^T H\right] \begin{bmatrix} \underline{d} \\ \underline{y}_r \end{bmatrix} \tag{3.37}$$

where

$$G_f = \begin{bmatrix} A & B \\ C & 0 \end{bmatrix} \text{ and } H = \begin{bmatrix} E & 0 \\ 0 & -I \end{bmatrix}$$

For any change in $\underline{\bar{x}}$ (and $\underline{\bar{u}}$) induced by $\underline{d}$ and/or $\underline{y}_r$, we have

$$\underline{\tilde{x}} = \underline{x} - \underline{\bar{x}}$$

and

$$\underline{\tilde{u}} = \underline{u} - \underline{\bar{u}} \tag{3.38}$$

Retaining the same feedback law as in (3.27), we get

$$\tilde{\underline{u}} = K_1 \tilde{\underline{x}} \tag{3.39}$$

after leaving out the integral of the output error term for the time being.

Substituting Eqs. (3.37) and (3.38) in Eq. (3.39), we get

$$\underline{u} = K_1 \underline{x} + [K_1 - I][G_f^T G_f]^{-1}[G_f^T H]\begin{bmatrix} \underline{d} \\ \underline{y}_r \end{bmatrix} \tag{3.40}$$

It is to be noted that effective feedforward control requires exact knowledge of the system model, that is, A, B, C, and E. For this reason, in practice, feedforward control is always used in conjunction with integral feedback control. The total control law thus becomes

$$\underline{u} = K_1 \underline{x} + K_2 \int (\underline{y} - \underline{y}_r) dt + K_{FF} \begin{bmatrix} \underline{d} \\ \underline{y}_r \end{bmatrix} \tag{3.41}$$

where

$$K_{FF} = [K_1 - I][G_f^T G_f]^{-1}[G_f^T H]$$

If K_{FF} is partitioned into K_3 and K_4, the feedforward term can be written as

$$\underline{u}_{FF} = K_3 \underline{d} + K_4 \underline{y}_r \tag{3.42}$$

3.6.3 Software for the Controller

The two main considerations in software development are:

(1) discretization of the continuous control and observer equations, sampling interval T being an important parameter, and

(2) approximation of the equation coefficients by integer values, as it is much simpler and faster to implement integer arithmetic in the processor.

The two choices for T are 5 ms for the thyristor chopper and 10 ms for the thyristor bridge rectifier. For an electromechanical system, faster action does not contribute much to the efficacy of the control. The approximation involved due to induction of integral arithmetic does not affect the quality of control to any great extent.

The software developed consists of the main program for computing the control and observer variables; integrations are changed to summations, negative numbers are represented in two's complement and a multiplication routine is used. The software package is sufficiently general and can be used for similar or more complex control laws without any difficulty. The supporting subroutines are for data acquisition and power line synchronization as described in Section 3.5.2. The time delay is implemented through the programmable timer/counter chip in the case of thyristor chopper as well as in the case of thyristor bridge rectifier, thus keeping the CPU (processor) free. The interrupt feature, though not used in this particular case, can be used for abnormal condition monitoring. The program in this case is written in Assembly language, though higher level languages such as Turbo-C or PROM-based object modules could be other options either for 16 or for higher bit processors/Personal Computers (PCs) or microcontrollers for low cost systems.

3.6.4 DC Motor Speed Control System

A separately excited, armature voltage controlled, dc motor is connected to an identical dc generator with lamp (resistive) load, in the laboratory.

The specifications of the dc machine are:

0.75 kW, 220 V, 4.8 A, 1500 rpm

The measured values of machine parameters and constants are:

$$R_a = 2.3\ \Omega, \qquad L_a = 46.5\ \text{mH}, \qquad k = 1.1\ \text{V}\cdot\text{s/rad or N}\cdot\text{m/A}$$

The field current was kept constant for both motor and generator. The combined load inertia, and friction of the motor-generator set, were determined by experiments and the values are:

$$J = 0.07\ \text{kg}\cdot\text{m}^2, \qquad \nu = 0.002\ \text{N}\cdot\text{m}\cdot\text{s/rad}$$

The speed of the set was picked up by a dc tachogenerator, its constant being 33.8 V · s/rad, and low pass filtered. The motor armature current was measured across a 0.1 Ω resistor placed in the circuit, amplified by an OP AMP of gain 10 and low pass filtered. The generator current, which is approximately proportional to load torque, neglecting losses, was measured across a 1.0 Ω resistor placed in the circuit.

If the brain of the speed control system is a programmable controller, the heart of the system is a power electronic controller such as the thyristor chopper or single-phase thyristor bridge rectifier. In the absence of a GTO, the class C chopper configuration is used. The chopper has been modelled as a linear gain element in spite of discontinuous conduction. The single-phase thyristor bridge rectifier has a nonlinear input-output relationship; linearization has been performed using a look-up table stored in memory.

DC motor model

The electrical port equation of the motor is

$$L_a i_a + R_a i_a + k\omega = v_a \tag{3.43}$$

where

ω is the motor speed in rad/s, and

i_a and v_a are the armature current and voltage, respectively.

The mechanical port equation of the motor is

$$J\dot{\omega} + \nu\omega + T_L = ki_a \tag{3.44}$$

where T_L is the load torque in N · m.

With $[\omega\ i_a]^T$ as state vector, v_a as control input u, T_L as disturbance input d, and ω as output y, the state space model of the dc motor is

$$\underline{x} = \begin{bmatrix} -\frac{\nu}{J} & \frac{k}{J} \\ -\frac{k}{L_a} & -\frac{R_a}{L_a} \end{bmatrix} \underline{x} + \begin{bmatrix} 0 \\ \frac{1}{L_a} \end{bmatrix} u + \begin{bmatrix} -\frac{1}{J} \\ 0 \end{bmatrix} d \tag{3.45}$$

$$y = [1\ 0]\underline{x} \tag{3.46}$$

Determination of control law and observer equation

For a set point (reference) regulation to $\underline{y}_r$, the augmented system equation (3.17), after substituting numerical values, is

$$\dot{\underline{z}} = \begin{bmatrix} -0.0286 & 15.714 & 0.0 \\ -23.656 & -48.387 & 0.0 \\ 1.0 & 0.0 & 0.0 \end{bmatrix} \underline{z} + \begin{bmatrix} 0.0 \\ 21.505 \\ 0.0 \end{bmatrix} \tag{3.47}$$

The eigenvalues (poles) are $\lambda_1 = 0.0$, $\lambda_2 = -9.617$, and $\lambda_3 = -38.8$, of which the first one is the regulation pole and the other two are system poles. The system passes the controllability tests of Eq. (3.26). As the system poles are in the left half plane, only the regulation pole λ_1 is shifted to $\lambda'_1 = -20.0$ and the corresponding scalar gain k_1 of Eq. (3.22) is -22.082.

The integral feedback control law is

$$u = [-2.864 \;\; -0.93] \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} - 22.082 \int (y - y_r)dt \tag{3.48}$$

If feedforward control is added, the total control law is

$$u = -2.864x_1 - 0.93x_2 - 22.082 \int (y - y_r)\,dt + 2.89d + 3.97y_r \tag{3.49}$$

Reduced order observer

If instead of measuring the armature current directly, it is estimated by a reduced order observer as

$$\dot{\hat{\xi}} = -50.0\hat{\xi} - 28.784y + 21.505u + 1.466d \tag{3.50}$$

$$\hat{x}_2 = \hat{\xi} + 0.103\,y$$

where $d_1 = -50.0$ for a 'fast' observer and $f = -28.784$.

For combined observation and control, x_2 of Eq. (3.47) is replaced by $\hat{x}_2$ of Eq (3.50).

The discretized version of Eq. (3.50) with sampling time of 5 ms for processor implementation is

$$\hat{\xi}(k+1) = 0.7788\hat{\xi}(k) + 0.095u(k) + 0.0065d(k) - 0.1273y(k)$$

$$\hat{x}_2(k) = \hat{\xi}(k) + 0.103y(k) \tag{3.51}$$

Converted to integral coefficients, Eq. (3.51) becomes

$$\hat{\xi}(k+1) = \frac{199}{256}\hat{\xi}(k) + \frac{24}{256}u(k) + \frac{3}{256}v_3(k) - 4v_2(k)$$

$$\hat{x}_2(k) = \hat{\xi}(k) + \frac{7}{2}v_2(k) \tag{3.52}$$

Taking the chopper gain as 1.2, that is, output voltage to time delay, u is converted to T_{ON}, and the control law (3.49) for computation becomes

$$T_{ON} = -81v_1 - v_2 - 3\sum_{n=0}^{p}(v_1 - v_r) + 112v_r + \frac{5}{2}v_3 \tag{3.53}$$

Comments

State feedback controllers including observers have been developed for two types of ac motors—induction and synchronous. Reduced order observers are designed to estimate rotor and damper currents in induction and synchronous motors, respectively. These have also been implemented for induction motors using Intel 8085 (8-bit) microprocessor and nearly identical types of modules and for synchronous motors using Personal Computers (PCs) based on 16-bit processors. In the latter case, the standard Analog Input/Output (I/O) and Timer cards housed in PC along with interfacing circuits have been used. The ac motor model is nonlinear having terms proportional to speed, being one of the states. The system model has been linearized around an operating point and then the linear system theory is applied to determine the controller gains, which vary with the change in operating point. In some recent paper, the system has been divided into two subsystems—mechanical and electrical. A two-loop controller—speed controller, a standard proportional-cum-integral (P-I) type and current controller, state feedback type with a reduced order observer—has been developed. Also for the latter case, the nonlinear controller with speed as a variable component has been used to linearize the system. As these designs are complex, only the basic principle are stated here; the details are beyond the scope of the present text.

3.7 STEPPER MOTORS

An important aspect of the digital control of electric drives is the use of stepper motors. These motors are commonly used in computer peripherals, robotics, numerical controlled machine tools, x-y recorders, etc. A stepper motor operates on dc voltage in a digital manner on receipt of a command pulse or pulse train. It rotates in steps through a specific angle irrespective of whether it is off-load, or on-load, when pulsed. A brief introduction to stepper motors is presented below.

There are basically three types of stepper motors:

(a) Variable reluctance (VR) stepper motor

(b) Permanent magnet (PM) stepper motor

(c) Hybrid stepper motor

3.7.1 VR Stepper Motor

This motor works on the principle that a magnetic circuit always tends to maintain minimum reluctance for its flux path. There are only stator windings. The stator and rotor teeth provide variable reluctance. The stator windings are excited from a dc source in sequence. The stator windings of a four-phase, eight-pole VR stepper motor are shown in Fig. 3.25.

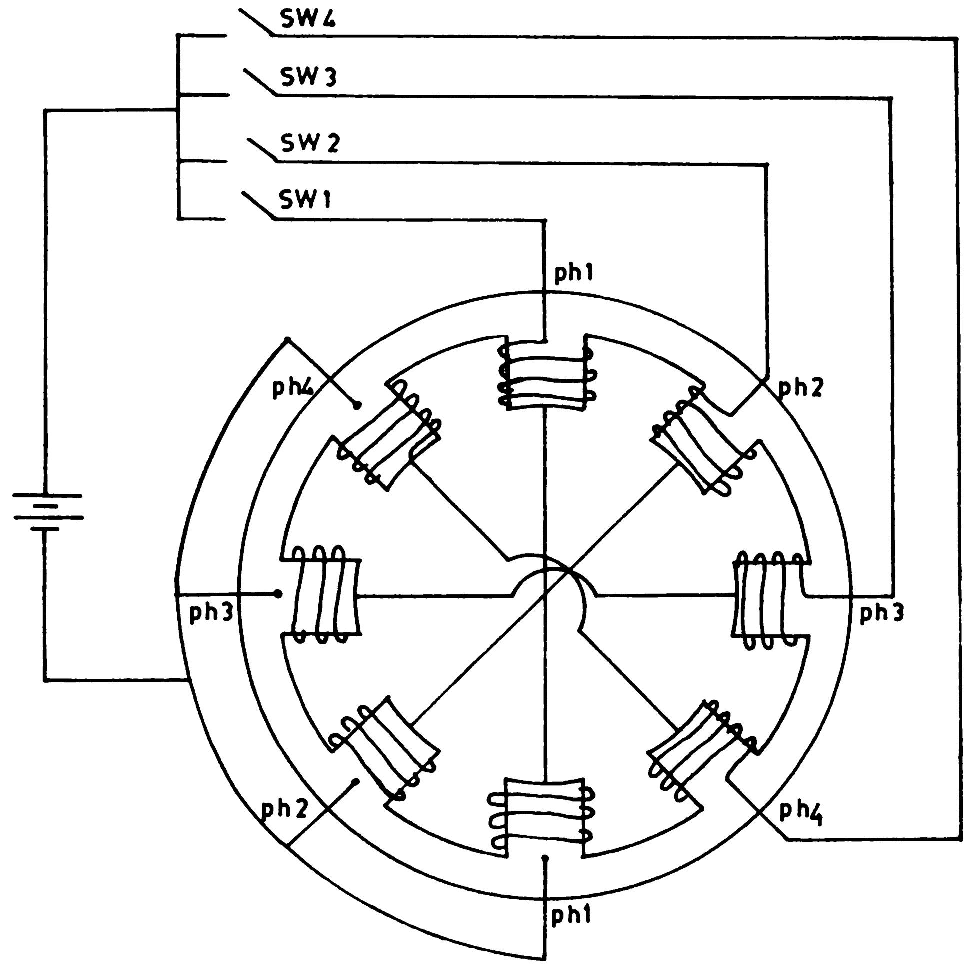


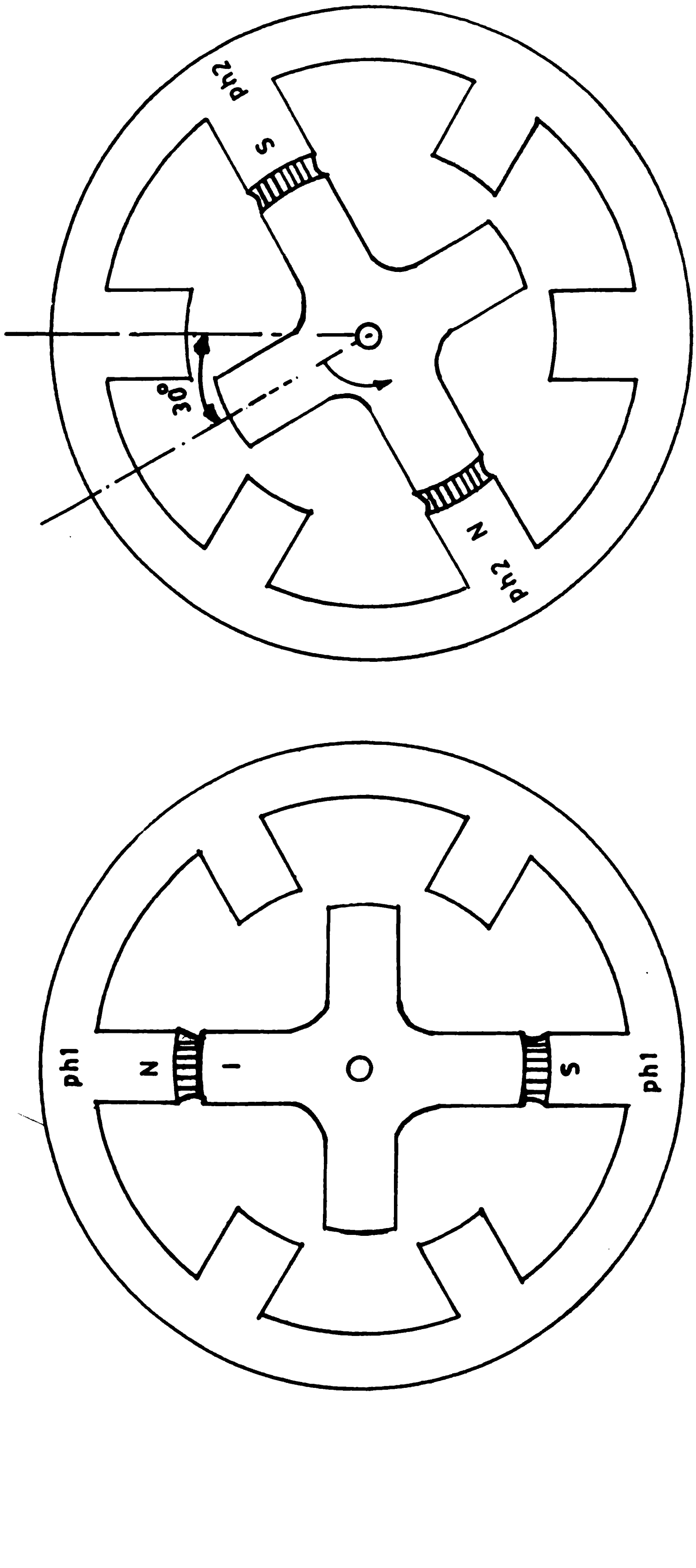
Fig. 3.25 Cross-sectional view of a four-phase, eight-pole VR stepper motor.

The number of coils for each phase winding may be one or more depending upon the number of stator teeth per phase. Six numbers of rotor teeth are required for a step angle of 15°. Similarly, there are six stator poles 60° apart for a three-phase motor. The motor rotates through 30° in each step, if the number of rotor teeth is equal to four. The operation of a three-phase stepper motor is illustrated in Fig. 3.26. When phase 1 is switched OFF and phase 2 is switched ON, the rotor rotates through a step angle of 30°. When phase 2 is switched OFF and phase 3 is switched ON, the rotor again jumps through 30°. In this way, the rotor rotates in steps from one position to another. The length of the air gap should be very small.

The step angle is given by

$$\theta_s = \frac{360°}{S} \tag{3.54}$$

$$S = mN_r \tag{3.55}$$



(a) Phase 1 is energised.

(b) Phase 2 is energised.

Fig. 3.26 Rotation in steps of a three-phase VR stepper motor.

where

θ_s = step angle

S = number of steps per revolution

m = number of phases

N_r = number of rotor teeth

The step angle is decreased by increasing the number of rotor teeth. For example, if $N_r = 50$ of a four-phase motor, the step angle is equal to 1.8°. The number of stator teeth is increased to increase the holding torque. The cross-sectional view of a four-phase stepper motor with 1.8° step angle is shown in Fig. 3.27.

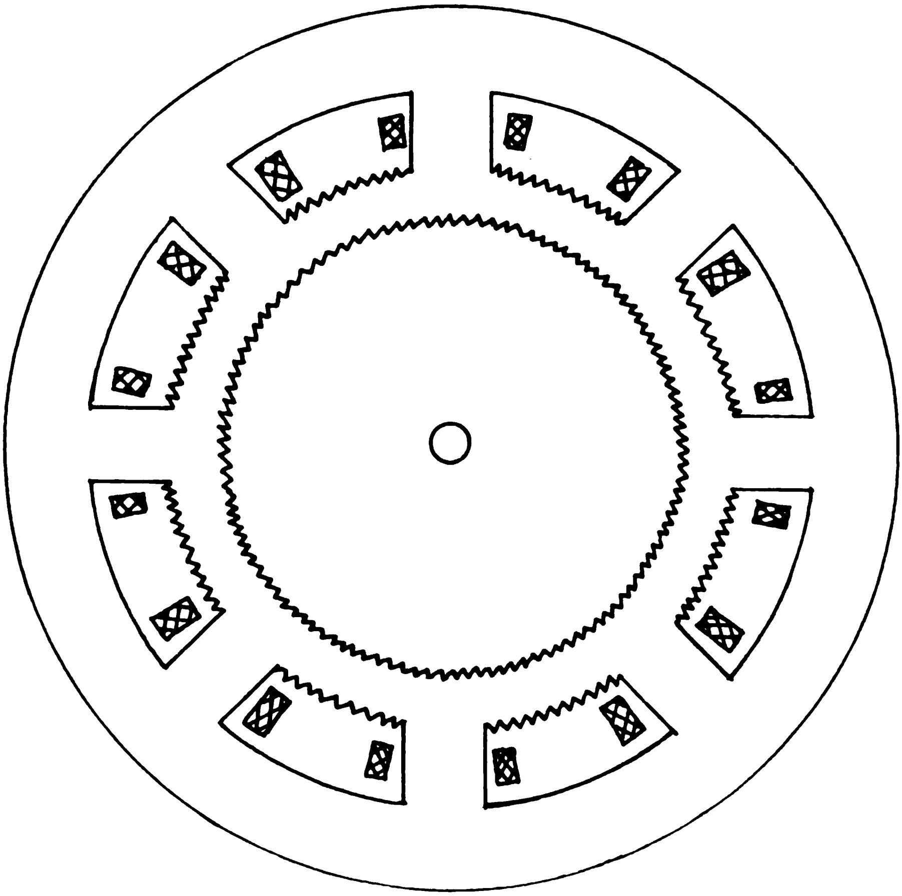


Fig. 3.27 Cross-sectional view of a four-phase VR stepper motor with 50 teeth.

The VR stepper motors may be constructed in single stack as described or in multistack in which stator and rotor portions are different for each phase. They are placed in cascade along the rotor shaft. The stator and rotor have the same tooth pitch in a stack but have different pitches for different stacks. Only one phase is excited at a time. There are different step angles for different stacks.

3.7.2 Permanent Magnet (PM) Stepper Motor

Phase windings are placed on the stator teeth as in the VR stepper motor. Permanent magnets are fitted to the rotor. The rotor becomes locked to a stationary position even when unexcited. This feature is known as detent mechanism. The VR motor does not have the detent ability. The operation of a PM stepper motor is illustrated in Fig. 3.28. The step angle is 90°, which is very large. Permanent magnets are also very costly. This problem has been overcome in the hybrid motor.

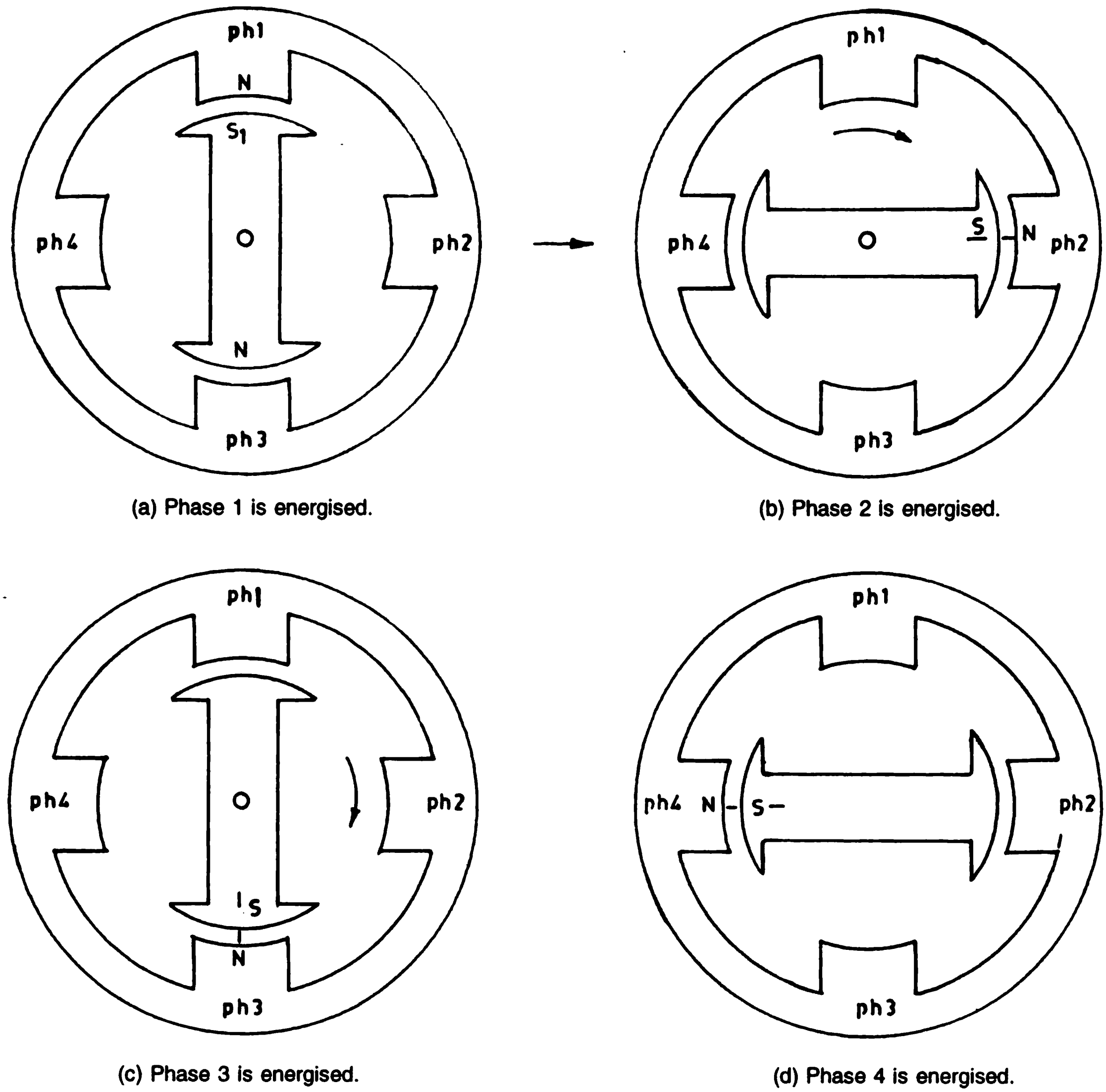


Fig. 3.28 Rotation of a four-phase PM stepper motor.

3.7.3 Hybrid Stepper Motor

In the hybrid stepper motor, a cylindrically shaped permanent magnet is placed in the rotor. The magnet is covered with soft steel with uniformly distributed teeth so that a small step angle is obtained.

3.7.4 Stepper Motor Control

The schematic of a four-phase stepper motor control is shown in Fig. 3.29. Proper fields are produced in the motor by the combination of switching the motor through sw_1, sw_2, sw_3 and sw_4 to the supply. Power transistors are used as the switches. The transistorized switching circuit is called the motor driver. When a pulse is fed to the base of the transistor, it conducts and the respective winding is energized. The pulses appearing at the bases of the transistors are generated by another circuit, called the logic sequencer. On receipt of a command pulse the logic sequencer determines the phase(s) to be energized or de-energized and sends pulses to the bases of the transistors accordingly.

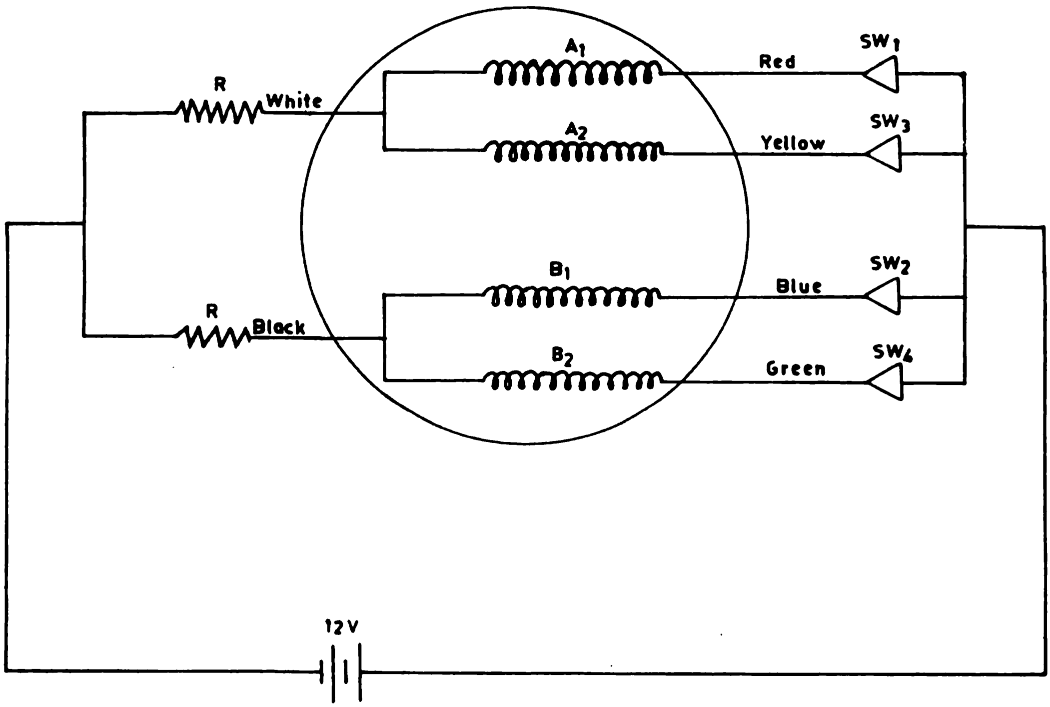


Fig. 3.29 Schematic of a four-phase stepper motor.

The sequence of a stepper motor is given by the manufacturer. The logic sequencer generates pulses in that order on receipt of the command pulses. The switching sequence can also be generated by a microprocessor. The block diagram of the motor drive system is shown in Fig. 3.30.

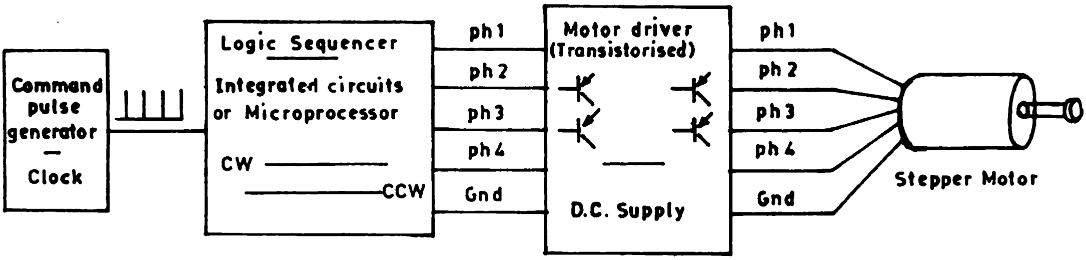


Fig. 3.30 Block diagram of the four-phase stepper motor drive system.

Command pulse generator

Command pulses can be generated, for example, by a dual monostable chip (74123). The novelty of this chip is that the operation of the chip can be started and stopped with the help of inhibit pin no 2. It can be used as a bistable multivibrator by connecting the monostables back-to-back.

Logic sequencer

Let us suppose that switching logic supplied by the manufacturer for the control of a four-phase stepper motor is given in Table 3.1. The steps (1–4) are given in sequence for rotation of the motor in the clockwise (CW) direction. The steps (4–1) in the reverse order are for rotation in the counterclockwise (CCW) direction. The steps are obtained using the 4-bit ring counter based on D-flip-flops, 7474 (Fig. 3.31a). The counter is loaded with 0011B. It is clocked once for each step.

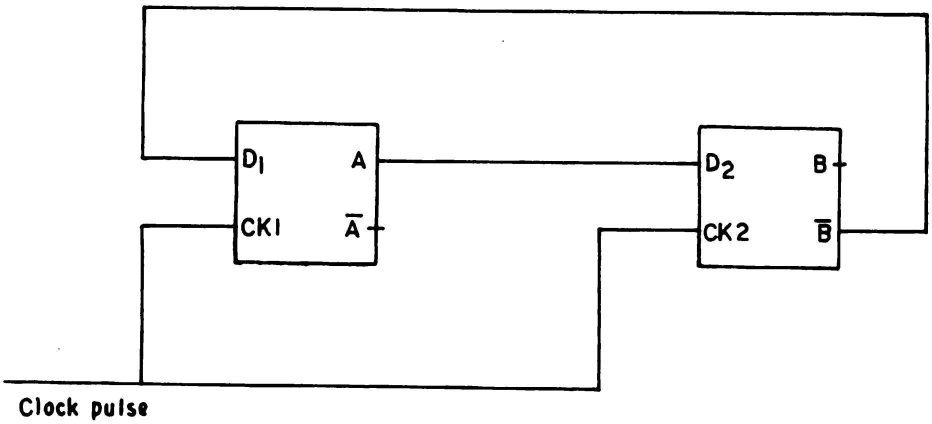
Table 3.1 Switching Logic

Step	*Switch*			
	$sw_4(B_2)$	$sw_3(A_2)$	$sw_2(B_1)$	$sw_1(A_1)$
1	0	0	1	1
2	1	0	0	1
3	1	1	0	0
4	0	1	1	0
5	0	0	1	1

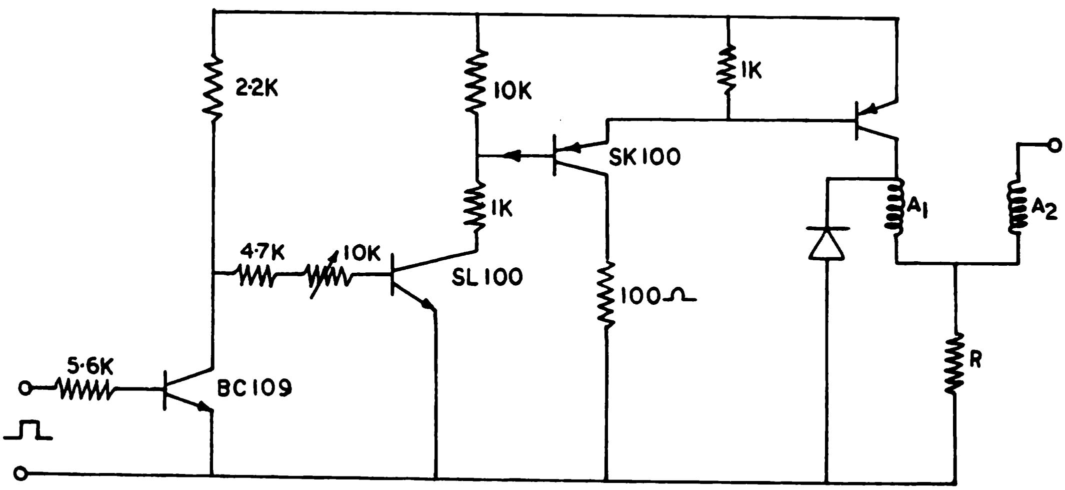
Another method to obtain the step sequence pulses is to use a microprocessor. The pulses may be generated at its output port, which in turn may be connected to the phase windings through the motor driver. 0011 0011B may be loaded into the accumulator. The logic sequence for clockwise rotation is obtained by rotating the byte right and outputting the four most significant bits. A short delay is introduced between the steps. Counterclockwise (CCW) rotation is obtained by rotating the byte to the left.

Motor driver

The output pulses from the 4-bit ring counter or the microprocessor are fed to the bases of the transistors in the preamplifier stage, whose output is amplified and then supplied to the motor windings. It may be noted that the amplifier circuits for A_1, A_2, B_1 and B_2 windings are exactly similar. A simplified diagram of the driver circuit is shown in Fig. 3.31b.



(a) Four-bit ring counter.



(b) Motor driver circuit for any of the phases.

Fig. 3.31 Circuits for the modules of Fig. 3.30.

PROBLEMS

3.1 A 220 V dc shunt motor takes 5 A and runs at 1000 rpm on no-load. The armature and field resistances are 0.2 Ω, and 220 Ω, respectively. Calculate the speed, developed torque and shaft torque, when the motor input is 11 kW.

3.2 A 220 V dc series motor takes 50 A, when giving its rated output at 1500 rpm. Its total resistance is 0.25 Ω. Determine the resistance to be added in series with the motor to obtain the rated torque at (a) starting, and (b) 1000 rpm.

3.3 A 230 V dc series motor used in lifts has a resistance of 0.25 Ω. It draws 40 A at a speed of 1500 rpm. Assume that the magnetization curve is a straight line between zero and 40 A, and the flux per pole at 60 A is 20% greater than that at 40 A. Determine the resistance to be added in series with the motor for

(a) a speed of 3000 rpm at the current of 15 A, and

(b) a speed of 1000 rpm at the current of 60 A.

3.4 Two identical 220 V, 13 A dc shunt machines are used in a Ward-Leonard system. The generator is driven at a constant speed of 1200 rpm and its field current is varied from 0.1 A and 1.3 A. The armature and the field resistances are 0.5 Ω and 220 Ω respectively. The no-load armature current may be assumed to be negligible. Find the speed range (a) for no-load operation, and (b) for full-load operation.

The open circuit characteristic of the above machines is given below:

Field current, A	0.1	0.5	1.3
Open circuit emf, V	42	206	251

3.5 The speed of a 3.8 kW, 220 V, 20 A, 1000 rpm dc shunt motor is to be controlled by means of an adjustable voltage system. The constant speed dc shunt generator of the system is rated for 7 kW, 230 V, 30 A, 1470 rpm. The armature resistance of the generator and the motor is equal to 0.5 Ω each. Determine

(a) the motor ideal no-load speed,

(b) the generator excitation voltages and no-load speeds of the motor for the conditions at which the motor will operate at rated current and the speeds of 750, 500 and 250 rpm, and

(c) the limits of speed control at rated motor excitation on condition that the load current may be allowed to rise to twice the rated value for a short interval of time.

3.6 A dc series motor drives an elevator load that requires a constant torque of 175 N · m. The dc supply voltage is 460 V and the combined resistance of the armature and series field winding is 1.0 Ω. The rotational losses and armature reaction are negligible.

(a) The speed of the elevator is controlled by a solid-state chopper. At 60% duty cycle ($\alpha = 0.6$) of the chopper, the motor current is 50 A. Determine the speed and horsepower output of the motor and efficiency of the system.

(b) The elevator is controlled by inserting resistance in series with the armature of the motor. For the same speed and output power as in part (a), determine the value of the series resistance to be inserted, horsepower output of the motor and efficiency of the system.

3.7 A series-wound fan motor runs normally at 600 rpm on 250 V dc supply, taking 20 A. The torque required by the fan increases as square of the speed. The motor resistance is 1.0 Ω. Assume that the flux is directly proportional to the current and ignore losses.

(a) Determine the power delivered to the fan and the torque developed by the motor.

(b) If the speed is reduced to 400 rpm by inserting a resistance in the armature circuit, determine the value of the resistance, the motor current and the power delivered to the fan.

3.8 A single-phase half-controlled bridge converter drives a 110 V, 3.73 kW, 1200 rpm, dc shunt (separately excited) motor. The ac supply is 120 V, 50 Hz. The thyristors are fired at delay angle, $\alpha = 45°$. The motor current is 15 A and is assumed to be ripple free. The armature resistance is neglected. Assume that the converter is lossless. For this condition:

(a) Draw the waveforms of v_s, v_0, i_0, i_{T1}, i_{D1}, and i_s.

(b) Determine the power drawn by the dc motor.

(c) Determine the supply volt-amperes and supply power factor.

(d) Determine the rms value of the thyristor current and diode current.

3.9 The speed of a 220 V, 3.73 kW, 1000 rpm dc shunt motor is controlled by a single-phase full-converter. The ac supply is 240 V, 50 Hz. A very large inductance is connected in series with the armature. Assume the motor and converter to be lossless. The motor emf constant is 1.9 V · s/rad. For a speed of 1000 rpm and rated motor current, determine

(a) the firing angle of the converter,

(b) the rms value of the supply current and thyristor current, and

(c) the supply power factor.

Also, draw the waveforms of supply voltage, supply current, converter output voltage and current.

3.10 A three-phase full-converter is used to control the speed of a 220 V, 3.73 kW, 1200 rpm, dc shunt motor. The ac supply is 240 V, 50 Hz. The motor emf constant is 1.7 V · s/rad. The armature resistance is 1.5 Ω. For $\alpha = 60°$, the motor speed is 800 rpm. Determine

(a) the average value of motor current, assuming it to be ripple free,

(b) the rms value of the thyristor current and supply line current, and

(c) the supply power factor.

3.11 The power circuit during regenerative braking of a subway car is shown in Fig. 3.32. The dc motor emf constant is 2.1 V · s/rad and the dc bus voltage (V_s) is 440 V. Assume the motor current to be ripple free. At a motor speed of 800 rpm and motor current of 250 A, determine

(a) the duty cycle α of the chopper for the operating condition, and

(b) the power fed back to the bus.

Also, draw the waveforms of v_0, i_0, and i_s for a particular value of duty cycle.

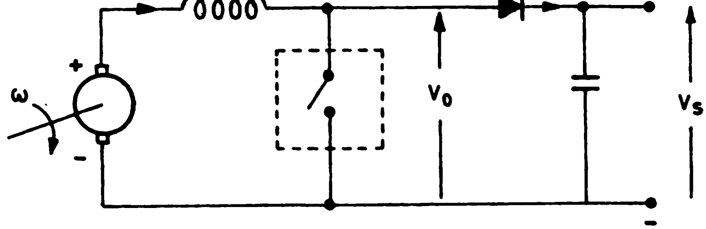


Fig. 3.32 Circuit for the Problem 3.11.

CHAPTER 4

SPEED CONTROL OF AC MOTORS

4.1 INTRODUCTION

Modern electric drives call for precise speed control over a wide range with long term stability and good dynamic characteristics. DC drives are still preferred, when precise speed control is required. AC drives are normally termed constant speed drives. But with the development of GTO/transistors (including MOSFETs) as high speed power switching devices along with control/data acquisition circuits based on microprocessors/Personal Computers (PCs), ac drives are now being used for variable speed applications.

There are certain disadvantages with dc drives. Regular maintenance is required for brushes and commutators, resulting in interruption of production schedule. DC drives are also not suitable for operation in a dirty and explosive environment. Further, maintenance problems may arise, if the drive is placed at an inaccessible place. Also, the dc motor has smaller power/weight ratio leading to its higher cost.

On the other hand, the induction motors, especially the squirrel cage type, are simple in construction, rugged in nature and require least maintenance. DC excitation is not required. Induction motors are the cheapest of all motors and are constructed with higher power ratings than those of dc motors. However, constant speed and low power factor are the main disadvantages. Until recently, the slip-ring induction motors and ac commutator motors constituted the only alternative to the dc drive, although unsatisfactory. The cost of power electronic converters sueh as inverters has now reduced considerably and their improved performance has resulted in increasing applications of ac motors to variable speed drives.

AC drives require variable voltage variable frequency power supply with constant Volt/Hz ratio to keep the flux constant in the motor. This results in constant torque operation at speeds below base (rated) speed. Above base speed, the input voltage is kept constant, resulting in constant horsepower operation. The flux in the motor decreases as frequency is increased.

Static Kramer drive is used for speed control of slip ring or wound rotor induction motors. Another control strategy employed is the use of phase locked loop (PLL). In this method, the error signal, i.e. the difference between the reference and output speed signals obtained by digital comparison, actuates a voltage controlled oscillator (VCO) so as to reduce the error, such that the output speed is locked with the reference speed. Other control schemes are also presented. For single-phase induction motors, suitable for low power applications, input voltage control is generally used, especially for fan type loads.

In the synchronous motor (wound field), dc excitation is normally provided by the rotor (field) circuit, which moves in synchronism with the rotating magnetic field, so that with fixed frequency supply, the motor operates at constant speed, corresponding to the supply frequency, i.e. synchronous speed. The main advantage is its ability to operate at different power factors such as lagging, unity or leading by controlling the field current as already described in Chapter 2. Permanent magnet and variable reluctance types of motors are now receiving attention for low and medium power applications. DC excitation is not needed in variable reluctance motors, torque being produced due to saliency in the rotor.

4.2 SPEED CONTROL OF INDUCTION MOTORS

The synchronous speed n_s of an ac motor is related to supply frequency f and pole pairs p by the equation

$$n_s = \frac{f}{p}$$

As regards induction motor, the rotor speed is given by

$$n_r = (1 - s)n_s$$

where s is the slip.

It is found from the above two equations that the basic methods of speed control of an induction motor are:

(a) by changing the number of poles and pole amplitude modulation technique, and

(b) by varying the line (input) frequency.

By the above two methods, the synchronous speed of an induction motor can only be changed. These methods are applicable to only cage induction motors.

The slip can be changed by the following methods:

(c) by varying the input voltage,

(d) by varying the rotor resistance,

(e) by cascade connection, and

(f) by injecting emf of slip frequency into the rotor circuit.

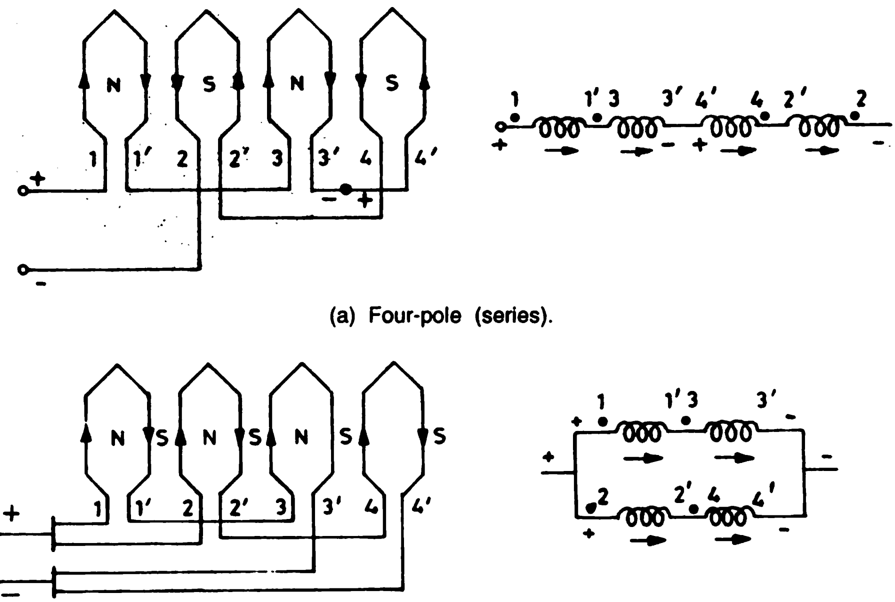
The methods (c–f) are applicable to slip ring (wound rotor) induction motors, whereas only the method (c) can be applied to machines with cage rotor. Speed control of induction motors has been described in all textbooks of electrical machines. Therefore, only some salient features are briefly presented here.

4.2.1 Pole Changing Induction Motor

Four-pole/eight-pole connections

In pole changing induction motors, the stator winding of each phase is divided into two equal groups of coils. These coil groups are connected in series and parallel with the current direction being reversed only in one group, to create two different numbers of poles (even) in the ratio 2:1 respectively. When the connection is changed from series to parallel or vice versa, the current in one group of coils is also reversed at the same time. This technique, termed the consequent pole method, is applied to all three windings (phases). This type of induction motor has always the squirrel cage rotor, which can adapt to any number of stator poles.

Figure 4.1a shows schematically only four coils of one phase of the windings connected in series, along with the direction of current in them, producing four poles in the stator. If the current in coils 2 and 4 is reversed and the connection is changed to parallel with two coils (1 and 3, and 2 and 4) connected in series for each path, eight poles are formed in the stator (Fig. 4.1b). It may be noted that the direction of current in coils 1 and 3 remains the same. Only one type ot connection is shown.



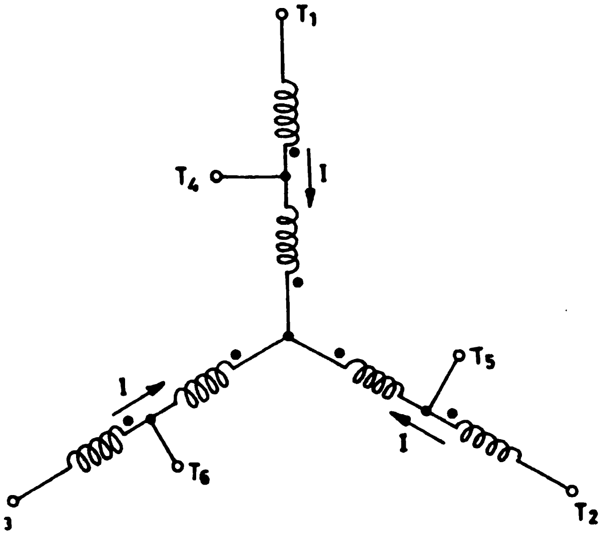
(b) Eight-pole (parallel) consequent pole.

Fig. 4.1 Stator winding connections for a pole changing induction motor.

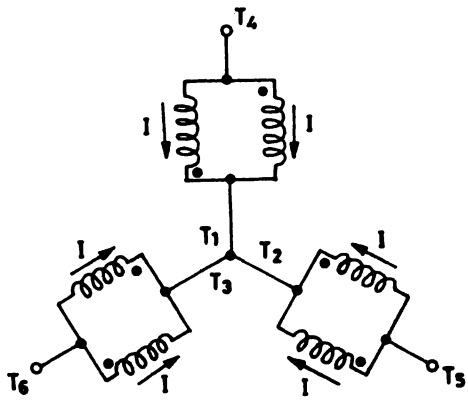
Constant torque and constant horsepower operations

This type of pole changing in the stator results in constant torque or constant horsepower operations. For constant torque operation, the change of stator winding is made from series-star to parallel-star, while for constant horsepower operation the change is made from series-delta to parallel-star. Regenerative braking takes place during changeover from higher to lower speeds.

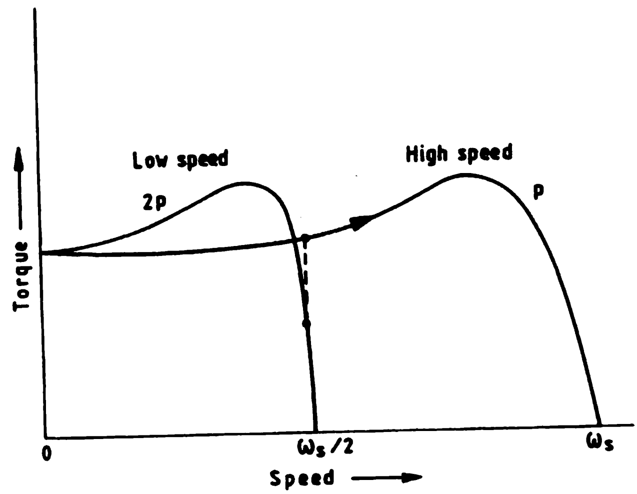
Constant torque operation. In any pole changing (p-pole/$2p$-pole) induction motor, there are two equal parts as stated above. The changeover for constant torque operation takes place as shown in Fig. 4.2.



(a) 2p-pole connection.



(b) p-pole connection.



(c) Torque-speed characteristics.

Fig. 4.2 Stator connections and torque-speed characteristics of induction motors for constant torque operation.

Let

V = Line voltage

I = Maximum current that the winding can carry

Then, the power drawn from the supply is given by:

1. For series-star (Y) connection (Fig. 4.2a),

$$P_Y = \sqrt{3}(VI \cos \phi_Y)$$

2. For parallel-star connection (Fig. 4.2b)

$$P_{YY} = 2\sqrt{3}(VI \cos \phi_{YY})$$

It is assumed that the power factor remains unchanged and the motor losses are negligible. With the changeover of stator winding from series star to parallel star, the power drawn from the supply is doubled. Simultaneously, the speed is also doubled. So, the motor torque remains constant. Constant torque operation is more common.

Constant horsepower operation. The power drawn from the supply is given by:

1. For series-delta (Δ) connection (Fig. 4.3a)

$$P_{\Delta} = 3(VI \cos \phi_{\Delta})$$

2. For parallel-star connection (Fig. 4.3b)

$$P_{YY} = 2\sqrt{3}(VI \cos \phi_{YY}) = 3.46\ (VI \cos \phi_{YY})$$

After changeover from series-delta to parallel-star, the power increases slightly (about 15%), if power factor is assumed to remain constant. The constant horsepower connection is the most expensive, because in this case the motor size becomes the largest.

4.2.2 Stator Frequency Variation

A variable frequency supply is considered as the key factor in speed control of induction motors. There are three basic types of variable frequency supply: (a) variable frequency motor-alternator set, (b) dc link inverter, and (c) cycloconverter. Nowadays, the variable frequency motor-alternator set is being largely replaced by static inverters of type (b). The inverter and cycloconverter are described in Sections 4.4 and 4.6, respectively.

Constant Volt/Hz operation

The emf per phase of an induction motor is given by

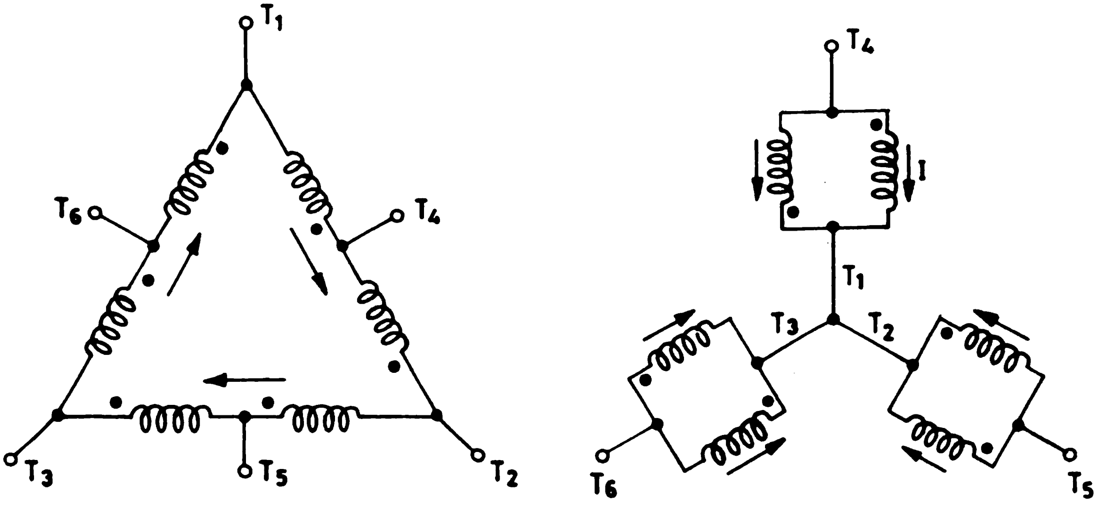
$$E = 4.44\ k_w \phi_m f T_{ph} \qquad \text{(V)} \tag{4.1}$$

with usual notations. The induced emf E is nearly equal to the applied voltage V (neglecting drop in stator impedance). Then, we can write

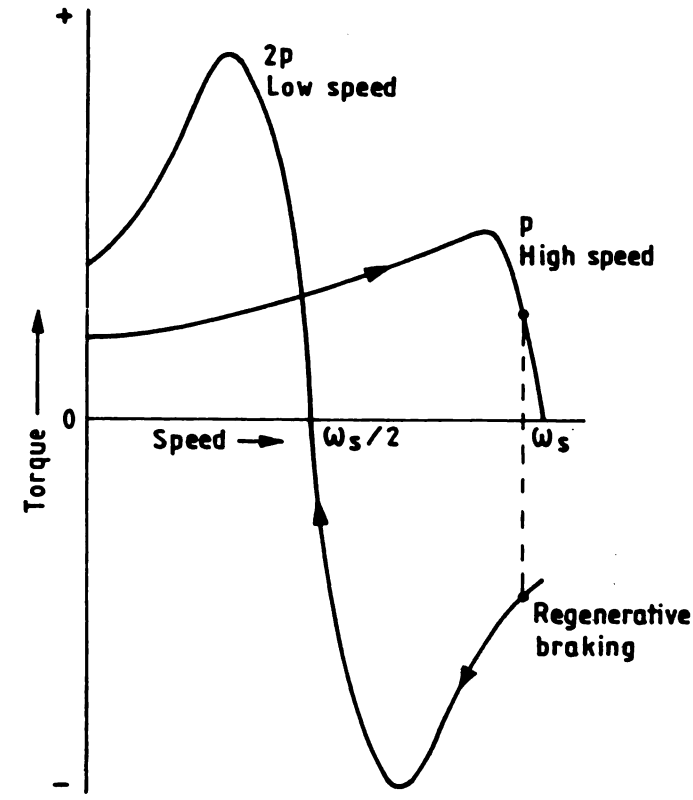
$$\frac{V}{f} = 4.44\ k_w \phi_m T_{ph} \qquad \text{(V/Hz)}$$

When the frequency is reduced, the applied voltage also must be reduced proportionally so as to maintain constant flux, otherwise the core will get saturated resulting in excessive iron loss and magnetizing current. The maximum torque also remains constant under this condition. However, the voltage is not varied proportionately in the low frequency range to account for the voltage drop in the winding resistance. This type of control (constant V/f) is used for speed control below base frequency (line frequency of 50 Hz).

As the voltage increases above rated value, when the input frequency goes above base frequency, only constant (rated) voltage with variable frequency (frequency control) is used for



(a) 2p-pole connection. (b) p-pole connection.



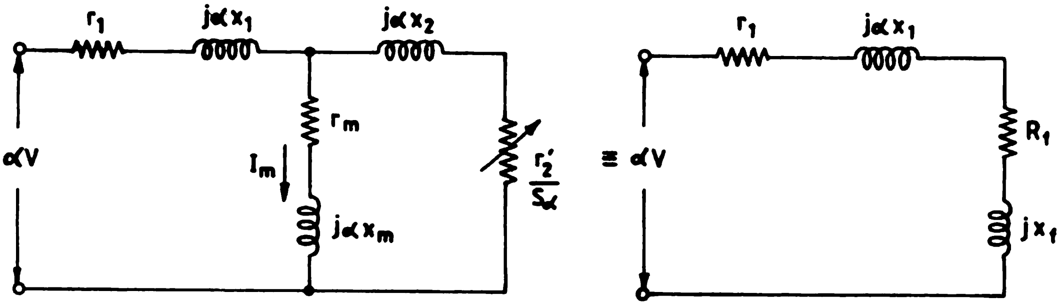
(c) Torque-speed characteristics.

Fig. 4.3 Stator connections and torque-speed characteristics of induction motors for constant horsepower operation.

speed control. Under this condition, both flux and maximum torque decrease as the frequency is increased.

Performance calculation

Steady state analysis of the induction motor under variable frequency supply can be based on the equivalent circuit shown in Fig. 4.4a.



(a) Actual (b) Simplified

Fig. 4.4 Equivalent circuit of the induction motor under variable frequency supply at constant air gap flux.

Here

$$s_\alpha = \frac{\alpha n_s - n_r}{\alpha n_s} \tag{4.2}$$

where

n_s = synchronous speed at base frequency

$$\alpha = \frac{\text{supply frequency}}{\text{base frequency}}$$

If R_m is neglected, it may be derived that

$$R_f = (\alpha X_m)^2 \frac{\dfrac{r_2'}{s_\alpha}}{\left(\dfrac{r_2'}{s_\alpha}\right)^2 + (\alpha x_{22})^2} \tag{4.3}$$

$$X_f = (\alpha X_m) \frac{\left(\dfrac{r_2'}{s_\alpha}\right)^2 + (\alpha^2 x_2' x_{22})}{\left(\dfrac{r_2'}{s_\alpha}\right)^2 + (\alpha x_{22})^2} \tag{4.4}$$

$$z = (r_1 + R_f) + j(\alpha x_1 + X_f)$$

The input current

$$I = \frac{\alpha V}{z} \tag{4.5}$$

The electromagnetic torque

$$T_e = \frac{3\alpha V^2 R_f}{\omega_s[(r_1 + R_f)^2 + (\alpha x_1 + X_f)^2]} \tag{4.6}$$

At starting, $s_\alpha = 1$, $\alpha x_{22} >> r_2'$ and $\alpha^2 x_2' x_{22} >> r_2'^2$.

So at starting, we have

$$R_f = (\alpha X_m)^2 \frac{r_2'}{(r_2')^2 + (\alpha x_{22})^2} \cong r_2' \tag{4.7}$$

$$X_f = (\alpha X_m) \frac{(r_2')^2 + (\alpha^2 x_2' x_{22})}{(r_2')^2 + (\alpha x_{22})^2} \cong \alpha x_2' \tag{4.8}$$

$$z_{st} = [(r_1 + r_2') + j\alpha\,(x_1 + x_2')] \tag{4.9}$$

For low values of α, the power factor improves at starting. The starting current is reduced. The starting torque is given by

$$T_{st} = \frac{3\alpha V^2 r_2'}{\omega_s[(r_1 + r_2')^2 + \alpha^2 (x_1 + x_2')^2]} \tag{4.10}$$

The maximum torque occurs when $dT_{st}/d\alpha = 0$ and $d^2T_{st}/d\alpha^2$ is negative.

$$\alpha_{max} = \frac{r_1 + r_2'}{x_1 + x_2'} \tag{4.11}$$

$$(T_{st})_{max} = \frac{3V^2 r_2'}{2\omega_s \alpha (x_1 + x_2')^2} \tag{4.12}$$

The advantages of constant Volt/Hz operation are the following:

(a) smooth speed control,

(b) small input current and improved power factor at low frequency start, and

(c) higher starting torque for low cage resistance.

Maximum torque

As the stator impedance is generally negligible compared to the input voltage, the equivalent circuit can be simplified to that shown in Fig. 4.4b. So, the equations can now be simplified as well. It may be noted that the input voltage is taken as (αV) with voltage to frequency V/f ratio being kept constant in this case.

The electromagnetic torque

$$T_e = \frac{3\alpha V^2\left(\dfrac{r_2'}{s_\alpha}\right)}{\omega_s\left[\left(r_1 + \dfrac{r_2'}{s_\alpha}\right)^2 + \alpha^2(x_1 + x_2')^2\right]} \tag{4.13}$$

The slip at maximum torque

$$s_{m\alpha} = \pm\frac{r_2'}{\sqrt{(r_1)^2 + \alpha^2(x_1 + x_2')^2}} \tag{4.14}$$

The maximum torque

$$T_{m\alpha} = \frac{3\alpha V^2}{2\omega_s(r_1 \pm \sqrt{(r_1)^2 + \alpha^2(x_1 + x_2')^2})}$$

$$= \frac{3V^2}{2\omega_s\left[\left(\dfrac{r_1}{\alpha}\right) \pm \sqrt{\left(\dfrac{r_1}{\alpha}\right)^2 + (x_1 + x_2')^2}\right]} \tag{4.15}$$

When the frequency is not low, $\left(\dfrac{r_1}{\alpha}\right) << (x_1 + x_2')$ and, therefore

$$T_{m\alpha} = \frac{3V^2}{2\omega_s(x_1 + x_2')} \tag{4.16}$$

So, the maximum torque remains constant as stated earlier for constant Volt/Hz ratio for frequencies below base frequency, except for very low values of frequency (Fig. 4.5). This is taken as constant torque control with constant flux or Volt/Hz ratio.

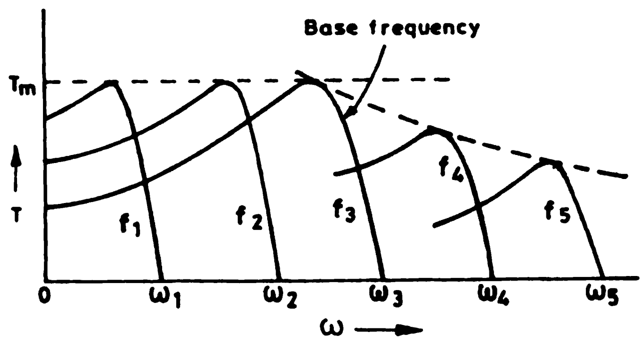


Fig. 4.5 Torque-speed characteristics of the induction motor under variable frequency supply.

Frequency control

For input frequencies above base frequency, only constant (input) voltage with variable frequency is applied as stated earlier. As the input voltage in the earlier case is taken as (αV) due to constant Volt/Hz ratio, the maximum torque (Eq. (4.16)) changes to

$$T_{m\alpha} = \frac{3}{2\omega_s(x_1 + x_2')}\left(\frac{V}{\alpha}\right)^2 \tag{4.17}$$

With $\alpha > 1$ as frequency is higher than base frequency, both maximum torque and flux, as given by V/f ratio, decrease as frequency increases (Fig. 4.5). This is taken as constant power control with variable flux.

Current-fed induction motor

With the advent of three-phase current source inverter using thyristors described in Section 4.4.2, the current-fed induction motor drive has gained importance. For a constant input current, the rotor current depends on the relative values of the magnetizing and rotor impedances.
The rotor current

$$I_2' = \frac{jx_m I_1}{\left(r_1 + \frac{r_2'}{s}\right) + j(x_m + x_1 + x_2')} \tag{4.18}$$

The developed torque

$$T_e = \frac{3r_2'(x_m I_1)^2}{s\omega_s\left[\left(r_1 + \frac{r_2'}{s}\right)^2 + (x_m + x_1 + x_2')^2\right]} \tag{4.19}$$

The current I_1 through r_1 and x_1 is constant. Also generally, x_m is much greater than r_2' and x_2', and hence these can be neglected.

Neglecting the values of r_1 and x_1, the slip at maximum torque

$$s_m = \pm\frac{r_2'}{x_m + x_2'} \tag{4.20}$$

The maximum torque

$$T_m = \frac{3(x_m I_1)^2}{2\omega_s(x_m + x_2')} \tag{4.21}$$

It may be noted that the maximum torque depends on the square of the current and is independent of frequency. The starting torque is low, as x_m is large compared to x_1 and x_2'. The starting current is low due to low values of flux. As the speed increases (with decrease in slip), the stator voltage rises and the torque increases. With the increase in speed, the flux and magnetizing current are also increased. The torque is controlled by variation of input current and slip (Fig. 4.6).

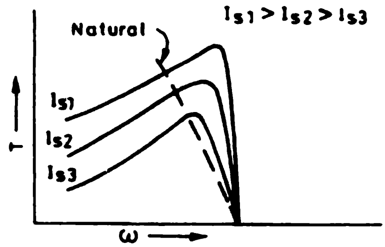
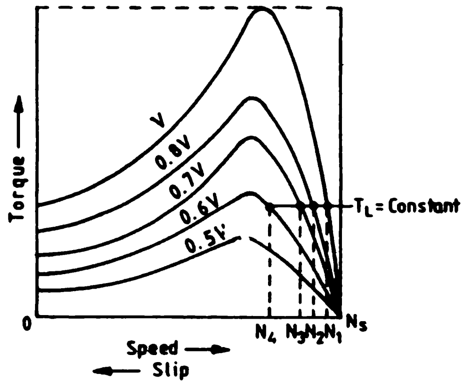


Fig. 4.6 Torque-speed characteristics of the induction motor supplied from constant current source.

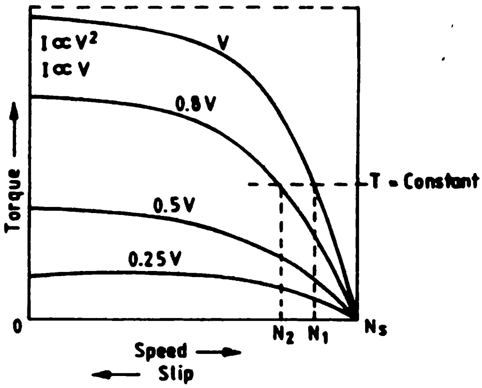
4.3 SPEED CONTROL OF SLIP-RING INDUCTION MOTORS

4.3.1 Stator Voltage Variation

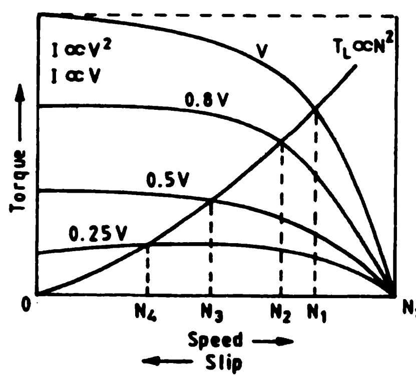
The developed torque of an induction motor at any slip is approximately proportional to the square of the applied voltage. The speed, therefore, can also be controlled by stator voltage variation. But for an ordinary induction motor operating with constant load torque, the range of speed control obtained is small (Fig. 4.7a). A continuous control of speed of an induction motor can be obtained



(a) Low rotor resistance, constant load torque.



(b) High rotor resistance, constant load torque.



(c) High rotor resistance, fan type load ($T_L \propto N^2$).

Fig. 4.7 Speed control of induction motors by variation of stator voltage.

by stepless adjustment of stator voltage, if its rotor resistance is high (Fig. 4.7b). The drawback here is that the torque per ampere is small at low speeds. However, in a fan or pump drive, the load torque varies as the square of the speed. The power requirement for these drives decreases rapidly with decrease in speed, thus surmounting this problem (Fig. 4.7c). Solid iron rotors may also be employed for satisfactory operation at low speeds. Otherwise, a piece of permanent magnet (e.g. Alnico) may be placed on the top of each slot to increase hysteresis loss and thereby the effective value of rotor resistance. The deep bar or double cage rotors are not suitable for this purpose.

4.3.2 Rotor Resistance Variation

Where the motors drive loads with intermittent type duty, such as cranes, ore or coal unloaders, skip hoists, mine hoists, lifts, etc. slip-ring induction motors with speed control by variation of resistance in the rotor circuit are frequently used.

The speed is controlled in steps. The same resistance can be used for starting the motor. High starting torque, low starting current and large pull-out torque are obtained. Excessive power loss at low speeds is one of its major shortcomings. The torque-speed characteristics (Fig. 4.8) loose hardness considerably at low speeds.

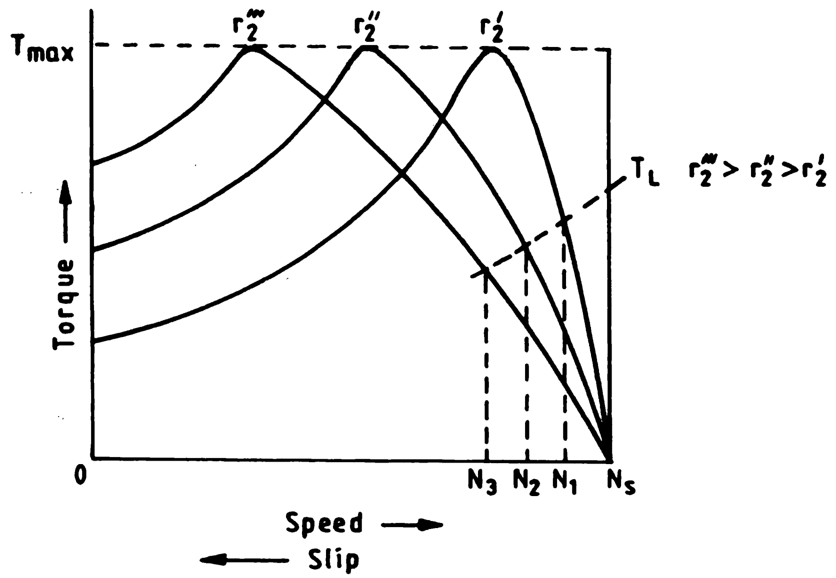


Fig. 4.8 Speed control of slip-ring induction motors by variation of rotor resistance.

The range of speed control is limited to 2:1 to 3:1. This method of speed control is employed for a motor-generator set with a flywheel (Ilgner set) used as an automatic slip regulator under shock loading conditions.

Rotor resistance control using a chopper

The rotor output terminals are connected to a three-phase diode (uncontrolled) bridge rectifier (converter), which is followed by a chopper (described in Section 3.5.2), with a GTO or transistor used as a switching device (Fig. 4.9). The inductor L_d is used in series with the chopper. The ON time of the chopper is varied so as to change the duty ratio α. Thus, the effective resistance is given by

$$R_e = R(1 - \alpha) \tag{4.22}$$

So, the speed can be controlled, if the duty ratio is changed. In this case, the slip power, which is the product of slip and power transferred via the air gap, is dissipated in the resistance R_e.

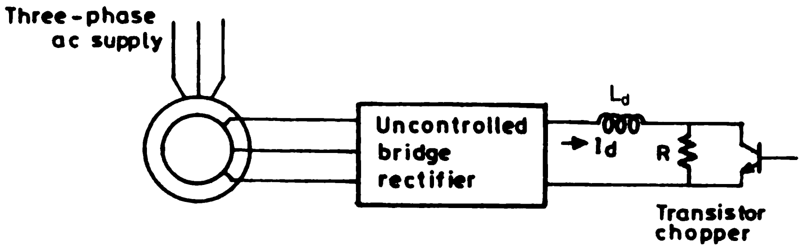


Fig. 4.9 Rotor resistance variation in slip ring induction motors using a chopper.

4.3.3 Slip Power Recovery

The phasor diagram (Fig. 4.10) shows the stator and rotor currents in an induction motor, when an emf of slip frequency is injected into its secondary circuit.

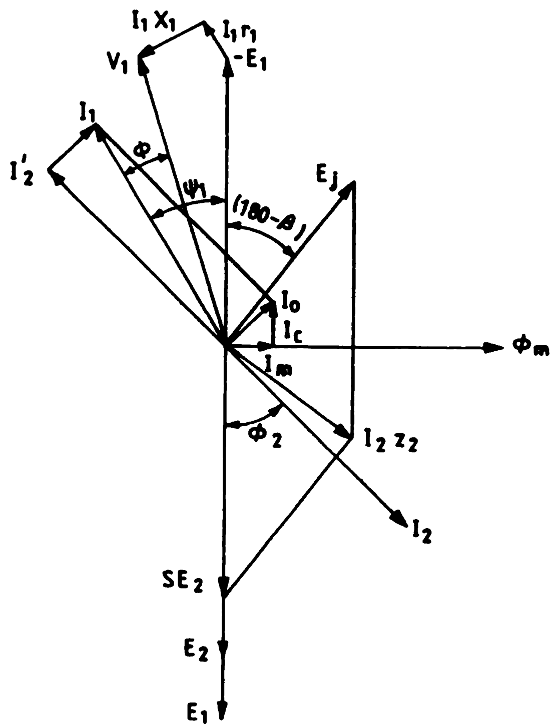


Fig. 4.10 Phasor diagram when an emf of slip frequency at an angle β is injected into the rotor circuit of the induction motor.

The power input to the induction motor is

$$P_1 = V_1 I_1 \cos\phi_1 = E_1 I_1 \cos\psi_1 + I_1^2 r_1 \tag{4.23}$$

Resolving currents on E_1, we have

$$I_1 \cos\psi_1 = I_c + I_2' \cos\phi_2$$

Multiplying by E_1, we get

$$E_1 I_1 \cos\psi_1 = E_1 I_c + E_1 I_2' \cos\phi_2$$

or

$$P_1 = I_1^2 r_1 + E_1 I_c + E_1 I_2' \cos\phi_2 \tag{4.24}$$

= (Stator copper loss) + (Iron loss) + (Power transferred to the secondary circuit)

Resolving emfs on I_2, we get

$$sE_2 \cos\phi_2 = I_2 r_2 + E_j \cos(180° - \beta + \phi_2)$$

Multiplying by I_2, we obtain

$$sE_2 I_2 \cos\phi_2 = I_2^2 r_2 + E_j I_2 \cos(180° - \beta + \phi_2)$$

Referring to the stator side, we have

$$sE_1 I_2' \cos\phi_2 = I_2'^2 r_2' + E_j' I_2' \cos(180° - \beta + \phi_2)$$

Adding $(1 - s)E_1 I_2' \cos\phi_2$, we get

$$E_1 I_2' \cos\phi_2 = I_2'^2 r_2' + E_j' I_2' \cos(180° - \beta + \phi_2) + (1 - s)E_1 I' \cos\phi_2 \tag{4.25}$$

Power transferred to the secondary circuit,

$$P_2 = \text{Rotor copper loss} + \text{Slip power} + \text{Mechanical power output}$$

The air gap power P_2 is nearly constant. If the losses are neglected, the mechanical power output and the slip can be varied, by regulating the slip power or injected voltage E_j at slip frequency sf.

Effect of injecting emf into the secondary circuit

When a slip frequency emf is injected into the secondary circuit in phase opposition to the standstill induced emf E_2 (Fig. 4.11a), the secondary current decreases resulting in reduced torque. The load torque is assumed to remain constant. The speed decreases, because the developed torque is less than the load torque. Then slip sE_2 and secondary current increase and as a result, the motor torque gradually increases until it balances the load torque. Under this condition

$$sE_2 = I_2 z_2 + E_j \quad (s = \text{positive}) \tag{4.26}$$

The angle of injection of emf β is equal to 180°. If $(I_2 z_2)$ is negligibly small

$$sE_2 = E_j$$

or

$$s = \frac{E_j}{E_2}$$

Hence, the speed can be controlled over a wide range below synchronous speed by varying the injected emf.

When the slip frequency emf is injected into the secondary circuit in phase with the standstill induced emf E_2 at $\beta = 0°$ (Fig. 4.11b), secondary current, motor torque, and speed increase as the load torque is assumed to remain constant. Under this condition, the emf equation is given by

$$I_2 z_2 = sE_2 + E_j \qquad (s = \text{negative}) \tag{4.27}$$

If $(I_2 z_2)$ is assumed to be negligibly small, we have

$$s = -\frac{E_j}{E_2}$$

Hence, the speed can be controlled above synchronous speed up to nearly double of the synchronous speed.

If the emf is injected into the rotor circuit at an angle $\beta = 90°$, there is no component of the injected emf directly assisting or opposing sE_2. Hence, the motor speed does not change. But the secondary current advances and the power factor increases due to the horizontal component of the injected emf (Fig. 4.11c). If the emf is injected at an angle $180° > \beta > 0°$, both the speed and power factor are controlled (Fig. 4.11d).

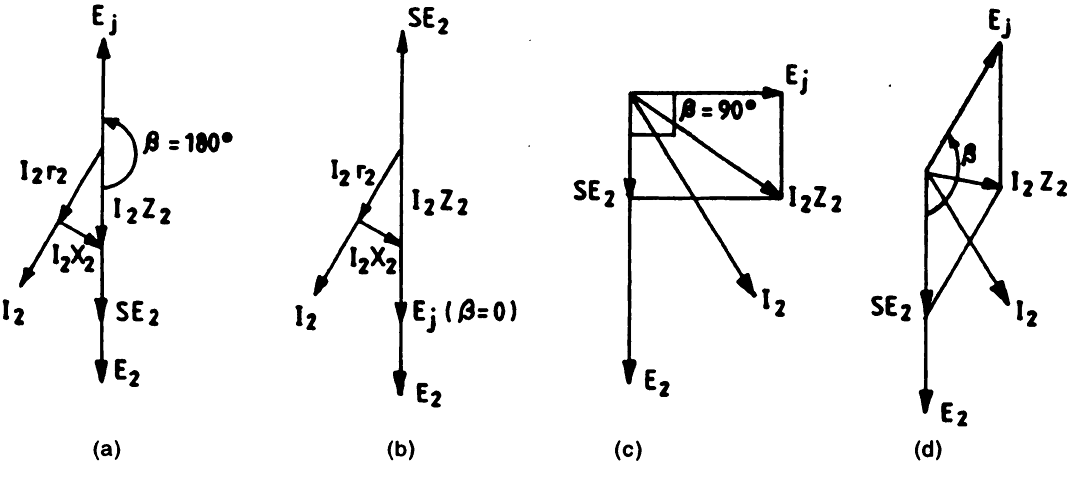


Fig. 4.11 Phasor diagrams for injected emfs at different β angles.

Kramer control

The static Kramer drive is shown in Fig. 4.12. The circuit consists of a three-phase diode bridge rectifier and a three-phase thyristor bridge rectifier (converter) acting as line inverter (described in Section 3.5.1), the output of which is fed back to the supply via a transformer. The rotor output at slip frequency (slip times line frequency) is converted to dc output and then converted to three-phase ac output at line frequency, using the bridge rectifier acting as inverter with delay angle, $\pi/2 < \alpha < \pi$. As its voltage output is low at low values of slip, a step-up transformer is used to increase the voltage to normal (rated) value.

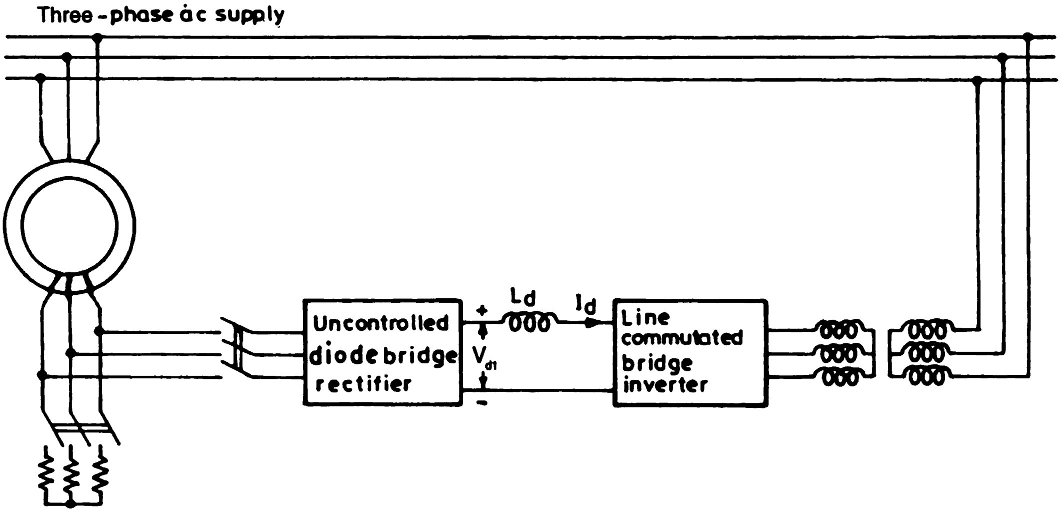


Fig. 4.12 Static Kramer drive.

The output voltage of the diode rectifier, neglecting stator and rotor voltage drops, is

$$V_{d1} = \frac{3\sqrt{6}}{\pi}\frac{sV}{n} \tag{4.28}$$

where n is stator to rotor turns ratio of the induction motor.

The output voltage of the thyristor converter (rectifier) acting as inverter is

$$V_{d2} = \frac{3\sqrt{6}\,V}{\pi}\frac{\cos\alpha}{m} \tag{4.29}$$

where m is source side to converter side turns ratio of the transformer.

As α is in the range $\pi/2 < \alpha < \pi$, V_{d2} is negative.

Neglecting voltage drops in the inductor connected in series, we have

$$V_{d1} + V_{d2} = 0 \tag{4.30}$$

From Eqs. (4.28 to 4.30), we get

$$s = -\frac{n}{m}\cos\alpha \tag{4.31}$$

The variation of delay angle permits power flow and speed control. In this case, the speed is only sub-synchronous (below synchronous speed). By changing the bridge rectifiers by three-phase dual converters or if both bridges are controlled rectifiers thus, permitting power flow in both directions, we can change the speed from sub-synchronous to super-synchronous (above synchronous) speed in induction motors.

Some continuous rolling mills, large air blowers, mine ventilators, centrifugal pumps and many other mechanisms including pump drives of hydraulic dredgers require speed adjustment in the range from 15 to 30% below or above normal. If the induction motor is of comparatively big size

(100 to 200 kW), it becomes uneconomical to adjust speed by means of external resistances due to copper losses as slip power is wasted as heat in the rotor circuit resistance. In these cases, the Kramer electrical drive system is used, where slip power recovery takes place. For speed control below synchronous speed, the slip power is pumped back to the supply, whereas for the case of speed above synchronous speed, additional slip power is injected into the rotor circuit. Kramer control schemes utilizing slip power recovery for constant torque and constant horsepower operations, have been illustrated in Fig. 4.13 (a and b) respectively.

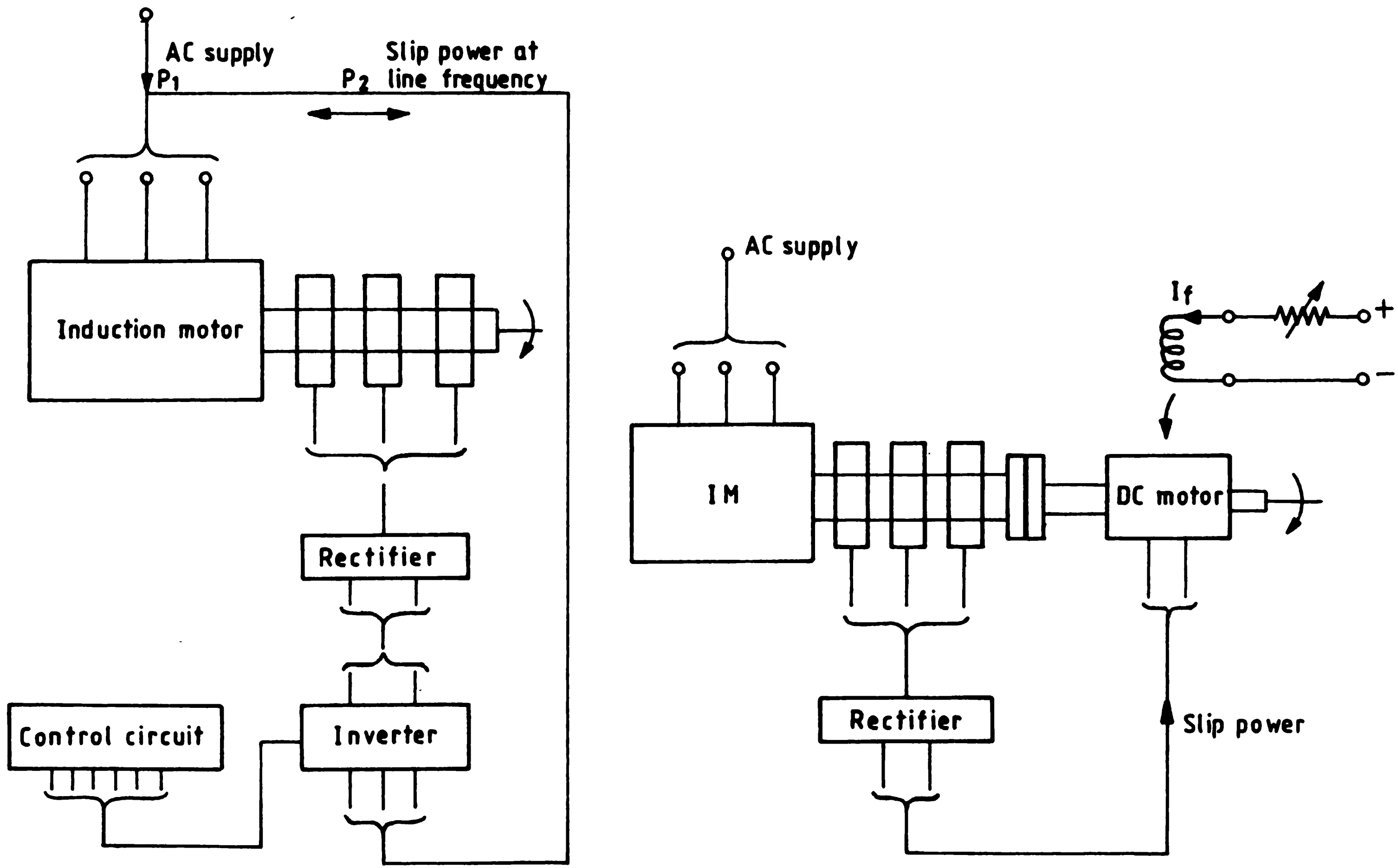
Vector control

This method is used for speed control of ac, both induction and synchronous, motors, though only the application to induction motors is briefly described here. In closed loop control, either stator voltage or stator current is regulated by the error in speed, but in magnitude only. The phase angle of the current with respect to flux is not considered. This is termed *scalar control*. But the stator current can be decomposed into two components in respect of rotor flux, the magnetizing and torque producing components, which are separately controlled. The transient response improves to a great extent using this method of vector or field-oriented control as this scheme resembles that of a normal dc motor in which field excitation and armature voltage are separately controlled.

The stator currents (phase) of the induction motor are decomposed into d- and q-axes components using a three-phase to two-phase transformation, with the rotor flux being available only along the d-axis, which is taken as the reference. The quadrature component of flux is zero. The system is now decoupled. Two controllers are used. One of them computes the magnetizing component from the error in flux between the reference value (as set or may be rated) and the measured/observed or computed one, while the other controller is used to determine the torque producing component from the error between the set value of torque and the measured value. The phase angle (θ) between the a-phase stator current and the rotor flux on the d-axis must be computed properly so as to determine the phase currents in the stator to be fed to the motor, from the two components using a two-phase to three-phase transformation. The dynamic performance of the closed loop system is superior to that of a system with scalar controller. The main problem is to either measure the flux directly using search coils, or estimate the flux indirectly using the motor model from terminal voltages and currents. The first one is known as the direct method, the second type is called the indirect method. In the direct method, flux observer has been used recently to estimate the rotor flux from the system (motor) model and use it to control the motor speed. This eliminates the need for search coils or Hall Effect sensors. In the indirect method, the variation of rotor resistance affects the machine model, though various methods have been proposed to compensate this effect. With the advent of different types of processors/PCs, it is now possible to design and then, implement a vector control scheme for speed control of ac (induction/synchronous) motors.

4.4 BASIC INVERTER CIRCUITS

The principle of inverter operation was known even before the advent of thyristors and conventional inverters were being fabricated earlier. With the development of high-speed thyristors and fast recovery diodes, various types of inverter circuits such as voltage source and current source inverters have been used for variable frequency operation of induction motors. Also, the use of microprocessors and other chips has greatly helped the development of sophisticated controllers for



(a) Constant torque drive.

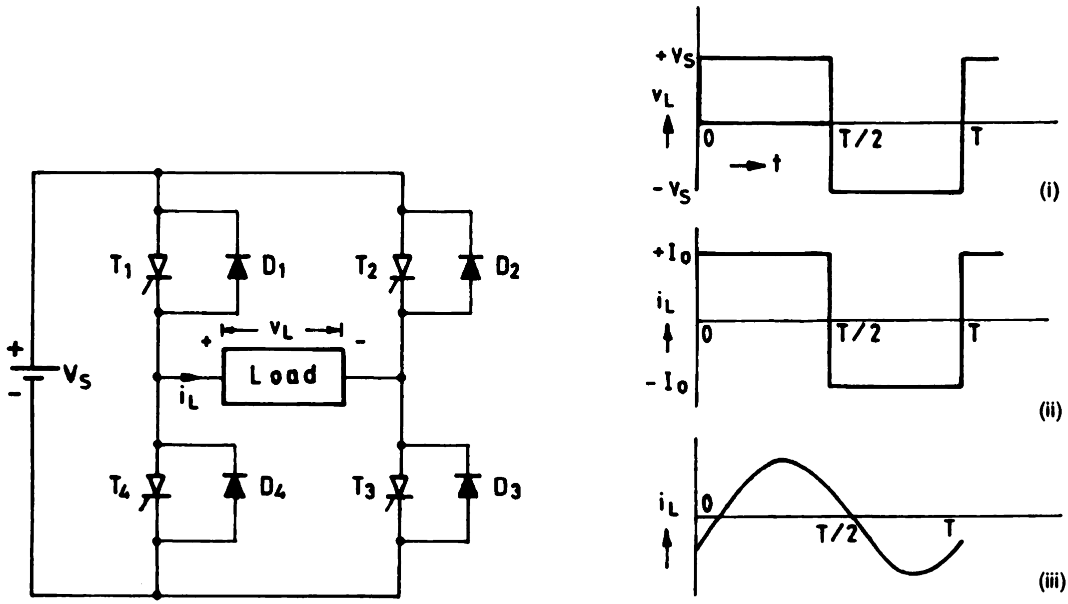
(b) Constant horsepower drive.

Fig. 4.13 Slip power recovery, Kramer control.

ac drives. As stated in Section 3.5, the availability of GTOs and transistors has not only influenced the inverter design, but also facilitated the use of Pulse Width Modulation (PWM) techniques due to their low switching time compared to that of inverter grade thyristors.

Single-phase bridge inverter

A single-phase bridge inverter (power circuit) is shown in Fig. 4.14a. The diodes are connected in parallel with thyristors in opposition. The voltage and current waveforms are shown in Fig. 4.14b. The voltage waveform is square, as the commutation is instantaneous. The current waveform follows that of voltage for resistive loads. The diodes are not needed for this type of load. But when the load changes to inductive type, the absence of diodes prevents proper functioning of the circuit as the commutation process of the thyristors is inhibited. But the diode(s) function during the time the current direction does not change, when the voltage is reversed across the inductive load. So, during this time the energy stored in the inductor is fed back to the source through the feedback diodes. The inverter frequency is controlled by proper firing signals at the thyristor gates.



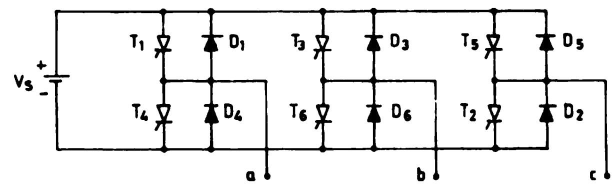
(a) Power circuit with reverse recovery diodes.

(b) (i) Voltage waveform; (ii) and (iii) Current waveforms for resistive, and inductive (R-L) loads, respectively.

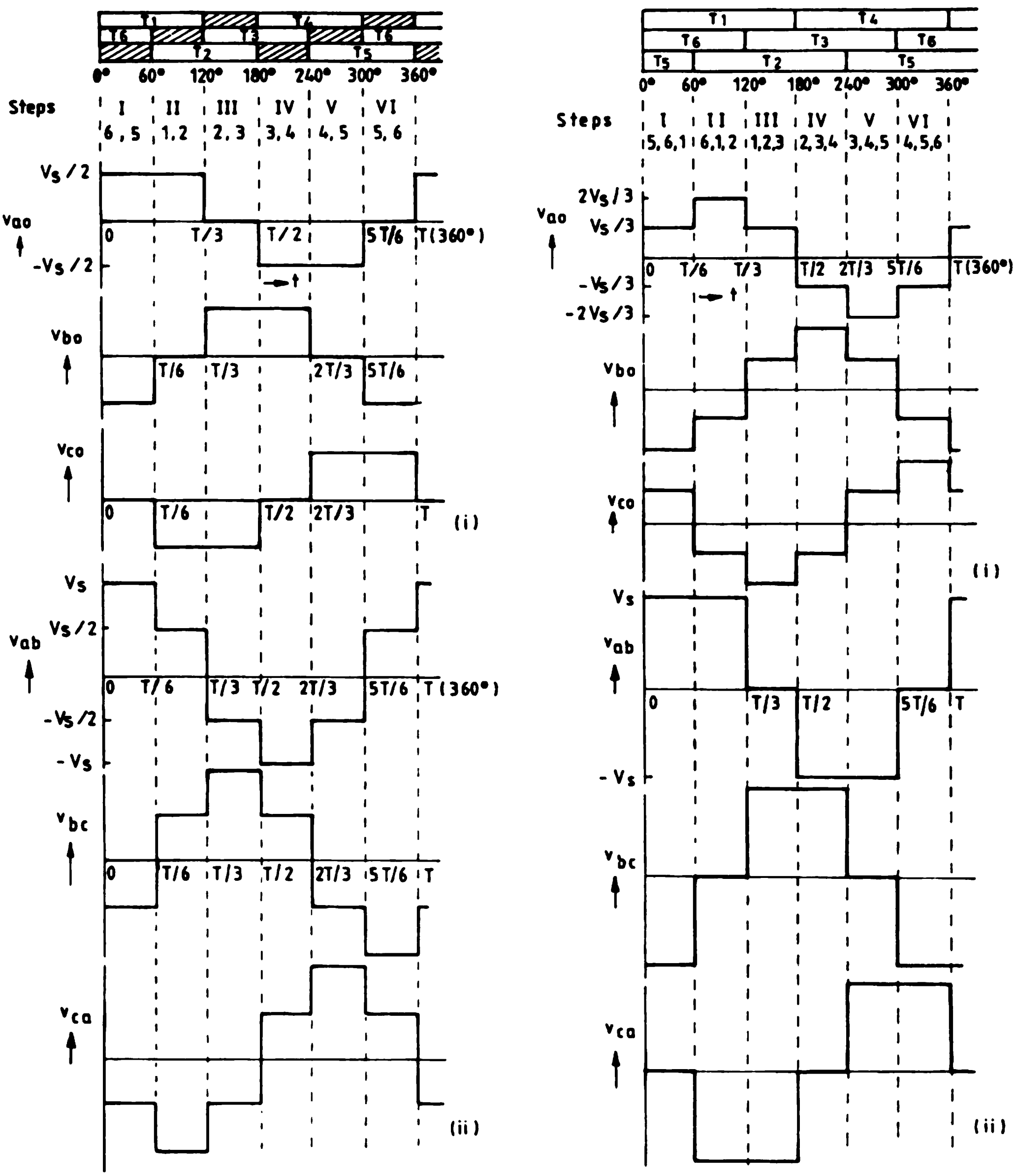
Fig. 4.14 Single-phase bridge inverter (voltage source).

Three-phase bridge inverter

A three-phase bridge inverter is shown in Fig. 4.15a with waveforms (Fig. 4.15b). For 120° conduction, the thyristors conduct in a sequence as shown in Section 3.5.1 for a diode/thyristor bridge ($\alpha = 0°$) as the commutation is assumed to be instantaneous. The feedback diodes are also to be used with inductive loads as described earlier. Only the power circuit is given here, the



(a) Power circuit with reverse recovery diodes.



(b) Six-step voltage waveforms for 120° mode conduction. (c) Six-step voltage waveforms for 180° mode conduction.

Fig. 4.15 Three-phase bridge inverter (voltage source).

commutation circuit is not shown. The frequency f is controlled in the same manner by firing the thyristors sequentially at intervals $T/6$ (60°), $T = 1/f$ being the time period. The line voltage waveform is stepped as shown.

Each thyristor can also conduct for 180° of a cycle. In this case, the thyristor T_1 conducts for first half (180°) of the cycle, while the thyristor T_4 conducts for second half (180°). The other thyristors are fired at an interval of 60°, as is the case above. The thyristors T_1, T_3 and T_5 in the upper group are fired at an interval of 120°, T_1 at $\theta\,(\omega t) = 0°$, T_3 at 120° and T_5 at 240°. The conducting sequence is shown in Fig. 4.15c. In each case for an interval of 60°, only three thyristors are conducting, one from the upper group and two from the lower group or two from the upper group and one from the lower group. For the first step ($60° > \theta > 0°$), the thyristors T_5, T_6 and T_1 conduct, in the second step ($120° > \theta > 60°$), T_6, T_1 and T_2 conduct and so on. The voltage waveforms are shown in Fig. 4.15c.

Commutation in inverters using modified McMurray-Bedford circuit

The circuit termed modified McMurray-Bedford type (Fig. 4.16) is used to commutate the conducting thyristor pair in a single-phase bridge inverter. The extra elements used are four capacitors and a pair of magnetically coupled inductors L_1 and L_4, and L_2 and L_3. The inductors are wound on a core with an air gap so as to avoid saturation. The capacitors are all equal. The auxiliary thyristors are not required. The commutation process is explained below briefly.

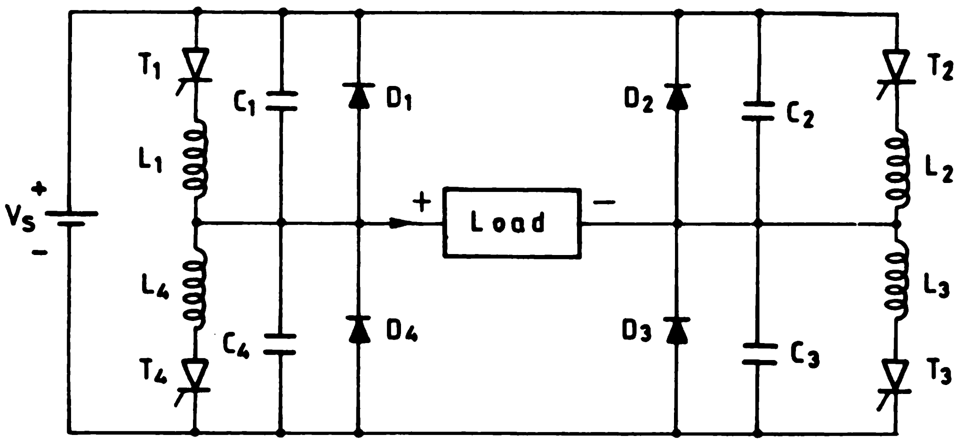


Fig. 4.16 Single-phase bridge inverter of modified McMurray–Bedford type.

The thyristor pair T_1 and T_3, is ON. The current is steady so that there are no voltage drops across L_1 and L_3. The two capacitors C_2 and C_4 are charged to voltage V through the respective conducting thyristors. The voltage across the other two capacitors C_1 and C_3 is nearly zero, as they are in parallel with the thyristors T_1 and T_3. At $t = 0^-$, the other pair of thyristors T_2 and T_4 is triggered to initiate the commutation of the conducting thyristors T_1 and T_3. As the voltages across the capacitors C_2 and C_4 cannot change instantaneously, the voltage V appears across L_2 and L_4, the equal voltage (V) is induced across L_1 and L_3, as the inductor pairs L_3 and L_2, and L_4 and L_1 are magnetically coupled. Thus, two conducting thyristors T_1 and T_3 are now subjected to the reverse voltage of $-V$, as the voltage across the two capacitors C_1 and C_3 is zero; the thyristors are turned OFF at

$t = 0^+$. The load current I passing through T_1, L_1, T_3 and L_3 is at once transferred to L_4 and L_2, so as to maintain constant mmf (proportional to I (L_1 or L_3) as the case may be) in the centre-tapped inductor as per the constant flux linkage theorem. It can be shown that the current through the two capacitor pairs C_1 and C_3, and C_2 and C_4 is the same as the load current I_0. Now, the capacitor pair C_1 and C_3 is getting charged and the other pair C_2 and C_4 is getting discharged. Also, the oscillating current flows through the paths C_4, L_4, C_2 and L_2, which rises to a maximum of I_m in T_4, L_4, T_2 and L_2 at the end of one-fourth of a cycle (t_1). At this moment (after t_1), the capacitor currents are $(I_0 + I_m)/2$.

At t_1, the capacitors, C_1 and C_3 are charged to voltage V and no current flows through the capacitors. At the same time the other capacitors C_2 and C_4 get completely discharged with the voltage coming to zero. Just after t_1, the current $(I_0 + I_m)/2$, through the capacitors C_2 and C_4 tends to charge them in the reverse direction and the diodes D_2 and D_4 get forward biased. Thus, the entire current, $(I_0 + I_m)$ is transferred to D_2 and D_4. So, the energy stored in the inductors (L_2/L_4) is dissipated in the closed circuit. The currents in the thyristors T_2 and T_4 decay from I_m at t_1 to zero at t_2 and the currents in the diodes D_2 and D_4 also decay from $(I_0 + I_m)$ at t_1 to some value i_0 at t_2. The thyristors T_2 and T_4 are OFF and the load current (i_0) flows through the diodes D_2 and D_4. This load current decays to zero at $t = t_3$ and is then reversed. As soon as the current tends to reverse, the diodes D_2 and D_4 are blocked. The diodes being now OFF, there is no reverse bias across the thyristors T_2 and T_4. At t_3, the thyristors start to conduct again due to the presence of gating pulses and carry the load current in the reverse direction. The capacitors C_1 and C_3, charged to voltage V are ready to commutate the main thyristors T_1 and T_3.

4.4.1 Variable Voltage Variable Frequency (VVVF) Control

The inverters described earlier are called 'dc link inverters', where the input dc voltage is obtained through either a fully controlled rectifier or an uncontrolled rectifier followed by dc a chopper, as described in Section 3.5. The controlled element is used to vary the input voltage to the inverter so that its output voltage may vary. A filter is always placed before the voltage source inverter to reduce the pulsation of the output voltage. For both induction as well as synchronous motors, variable frequency input requires that voltage to frequency ratio remains constant in order to maintain the flux constant. As the inverter frequency is varied, the voltage must be varied to maintain voltage to frequency (V/f) ratio constant. But this type of control requires two to three stages which results in higher losses and lower efficiency. Moreover, for low values of frequency, the elements of the filter circuit increase in size and weight which results in their high cost and low efficiency. Also for low values of dc input voltage, the voltage across the commutating capacitor decreases, thus reducing the circuit turn-off time for the thyristors for constant load current. The output voltage can also be varied by control within the inverter, such as (a) Series inverter control, and (b) Pulse Width Modulation (PWM) control. The first method involves the use of two or more inverters in series. For example, two inverters in series are used with their outputs fed to the load via transformers used for coupling. The voltage of one can be shifted in phase compared to the voltage of the other, such that the output voltage fed to the load can be varied, being the sum of the voltages of two inverters. The second method is described in the section below.

Pulse width modulated (PWM) inverter

The output voltage of this type of inverter is a number of pulses of varying periods in any cycle.

The ON and OFF periods are controlled in each half cycle by different modulation techniques. The output voltage is controlled, but the voltage to frequency (*V/f*) ratio is kept constant. Another advantage is that by proper control techniques, the lower order harmonics are reduced or eliminated, thus only the higher order harmonics need be filtered, which simplifies the problem. Only the devices, used in this case, must have low turn-on and turn-off times. With the advent of fast switching devices such as power transistors and MOSFETs, PWM techniques are being increasingly used in voltage source or current source inverters (discussed in Section 4.4.2). The modulation schemes used are: (a) multiple pulse, and (b) sinusoidal modulation. In the multiple pulse modulation method, the pulses are of equal width and the number of pulses per cycle is varied. In sinusoidal pulse width modulation methods, the crossover points to switch ON and OFF the devices are determined by comparing the sinusoidal reference waveforms with the triangular carrier waveform having frequency *n*-times the reference frequency (Fig. 4.17). The pulse widths are unequal, smaller at the beginning and end of each half cycle, increasing in width as the midpoint in each half cycle is gradually approached.

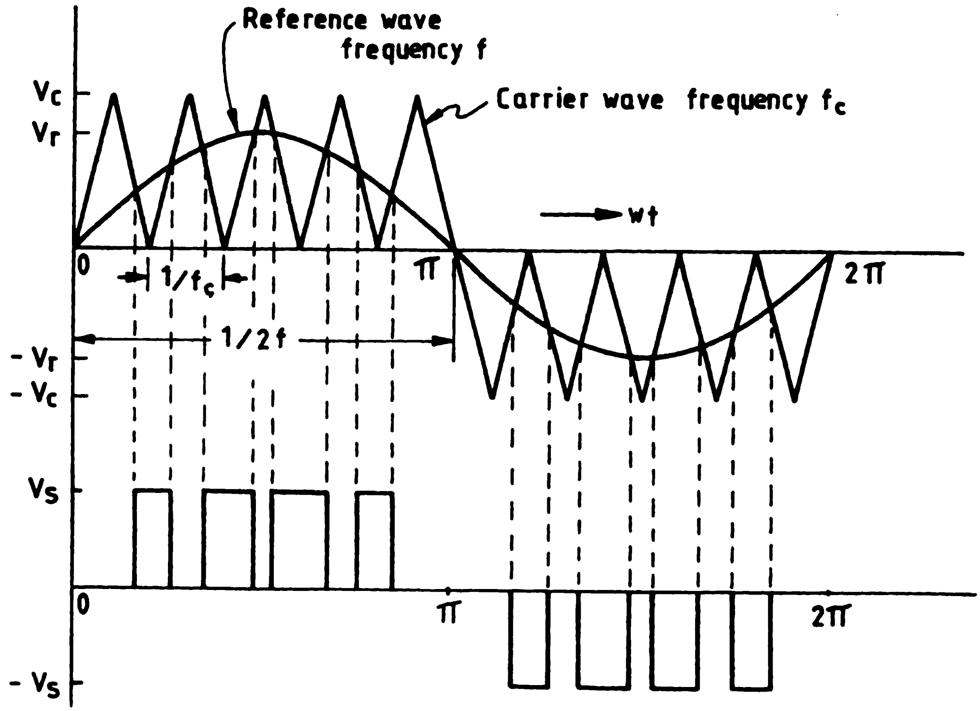
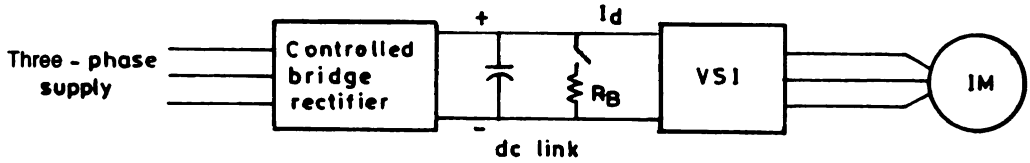


Fig. 4.17 Inverter voltage waveforms.

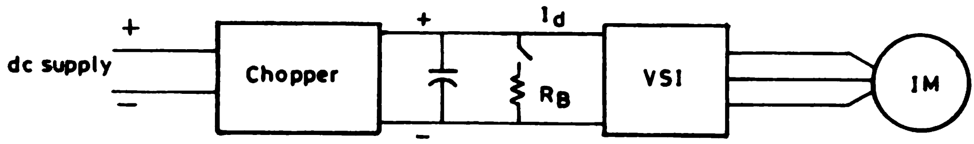
Voltage source inverter (VSI)-fed drive

The output waveform of a voltage source inverter can be either a simple stepped wave or a pulse width modulated wave. The latter type gives low harmonic content and, therefore, low pulsation of the torque, if the three-phase induction motor is fed from this type of inverter. For stepped waveform type, harmonics give rise to increased pulsation of the torque in an induction motor. The inverter frequency is controlled by the firing signals applied to the switching devices, such as transistors.

The input to such circuits is dc voltage, which is obtained either by full-controlled rectifiers (Fig. 4.18a) or by choppers (Fig. 4.18b), fed from ac and dc supply, respectively. In both these cases, the output voltage or dc input voltage to the inverter is varied so as to maintain Volts/Hz ratio constant in order to keep flux constant at rated value in ac motors. This method is used for starting as well as speed control of induction motors.



(a) Bridge rectifier.



(b) Chopper.

Fig. 4.18 DC link inverter (voltage source) fed induction motor.

For braking purposes, either the full-controlled converter needs to be replaced by a dual converter or the chopper by a two-quadrant chopper, in which case the power from the dc link inverter can be transferred to respective sources, to enable regenerative braking to take place. An inverter with variable frequency is used such that the frequency can be reduced so as to brake the induction motor. Dynamic braking is obtained by switching on a resistance R_b across the dc link. In this case, first the dc link voltage is sensed and as it exceeds a specified value, the resistance R_b is switched on to provide dynamic braking.

4.4.2 Current Source Inverter

In voltage source inverters, the output voltage is controlled by input voltage, which is kept constant, while the output current depends upon the type of load. In current source inverters (CSI), the input current is maintained constant, but it can be adjusted. For this purpose a high value of inductance is connected in series with the source to keep the current constant. The output voltage waveform is governed by load impedance. Though a three-phase CSI is needed to control ac machines, a single-phase CSI (Fig. 4.19) of auto-sequential commutation type is described here to state its basic principles.

The thyristor pairs T_1 and T_3, and T_2 and T_4 are alternately switched to obtain a nearly square wave load current. Two commutating capacitors C_1 and C_2, one in the upper half and the other in lower half are connected along with four diodes D_1 – D_4 These diodes connected in series with the respective thyristors prevent the capacitors from discharging into the load. For changing the inverter frequency, the thyristors must be triggered at suitable intervals ($T/2$), T being the time period. The

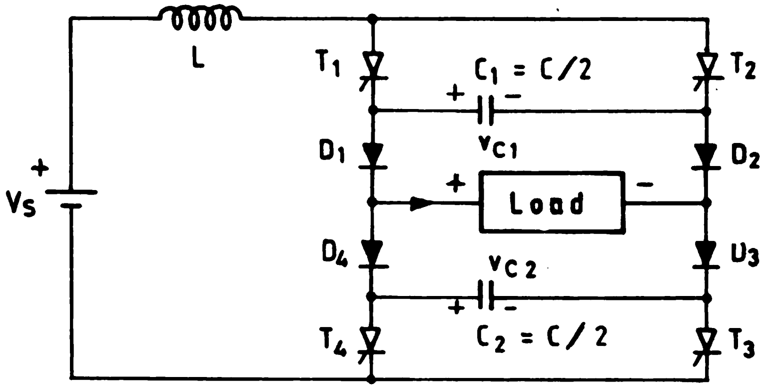


Fig. 4.19 Single-phase current source inverter (Auto-sequential commutation).

commutation process is explained. No additional components, except for diodes and capacitors, are needed. The frequency in this case is not high, but must be limited to allow the commutation process to be completed.

Mode I. It is initially assumed that the thyristor pair T_2 and T_4 is conducting, and current passes through the load and the diodes D_2 and D_4. The capacitors are charged to the same voltage, with the polarity such that the right hand plates are positive and the left hand plates negative.

The other thyristor pair T_1 and T_3 is now gated. The conducting thyristor pair T_2 and T_4 is turned off by the application of reverse voltages across the capacitors. The diodes D_2 and D_4 continue to conduct and both the capacitors are charged through the load and the diodes D_2 and D_4 and thyristors T_1 and T_3. As the direction of current shows, the capacitors are charged linearly by the load current, assumed to be constant. The voltages decrease. The other diodes D_1 and D_3 are initially reverse biased. But as the voltages across capacitors decrease with the polarity as shown earlier, the two diodes D_1 and D_3 are forward biased, when the capacitor voltages become zero and then the polarity reverses with the charging current as shown.

Mode II. Now all the four diodes $D_1 - D_2$ are conducting. The diodes D_1 and D_3 are forward biased with the capacitor voltages reducing to zero. The current passes through two paths. The load is now connected across the two capacitors in parallel. Due to the inductance in the load circuit, the current through the capacitors becomes zero after some time. This current also passes through the diodes D_2 and D_4. At the same time the current in the diodes D_1 and D_3 increases, thus reversing the load current. When the capacitor current reaches zero, the diodes D_2 and D_4 are OFF and the load current reaches the constant value in the reverse direction. The current path is now through T_1, D_1 load, D_3 and T_3. The capacitor voltage is now fully reversed with the polarity set so as to be ready for commutating the thyristor pair T_2 and T_4, when the pair is gated ON.

Current source inverter (CSI)-fed drive

This type of inverter is generally considered to be more reliable as the chance of commutation failure compared to a voltage source inverter is minimal. Regeneration is also possible. The presence of a large inductance L_d in the dc link makes the current constant. The torque is controlled by link current I_d, which is obtained by varying the input voltage V_d, using a full-controlled bridge rectifier or chopper as stated earlier. Normally, closed loop operation is required for CSI-fed drives. For regenerative braking, the machine runs as a generator, when the inverter frequency is reduced. The full-controlled converter replaced by a dual converter (as shown for VSI) works as an inverter and CSI acts as a converter feeding power to the dc link.

4.5 AC REGULATOR

The operation of ac regulators is mostly similar to that of phase-controlled converters. The main use of such regulators is for voltage control as used for heating and lighting and also speed control of mainly single-phase and three-phase induction motors (with fan type load) including starting.

A single-phase ac voltage controller or regulator is shown in Fig. 4.20. Two thyristors can be connected back-to-back or a triac may be used instead. A triac, unlike thyristors, can conduct in both directions, when gated. A triac goes off, when the current reaches zero, and only then it can be gated to conduct in either direction. The triac or thyristor T_1 is gated at a delay angle α from $\theta = 0°$ for conduction in the forward direction. When the voltage decreases to zero, the current decreases too, but does not become zero immediately, due to inductive load. It decreases to zero after some time, when the triac or thyristor T_1 goes off. The triac or the thyristor T_2 is gated at $\theta = \pi + \alpha$, with delay α after the voltage reaches zero. The conduction continues till the current decreases to zero as stated earlier. The current is now in reverse direction as the voltage is reversed after $\theta = \pi$. Thus, with change of delay angle α, the output voltage of the regulator changes too. Such controllers can be used for speed control of fan drives as load torque is proportional to square of speed.

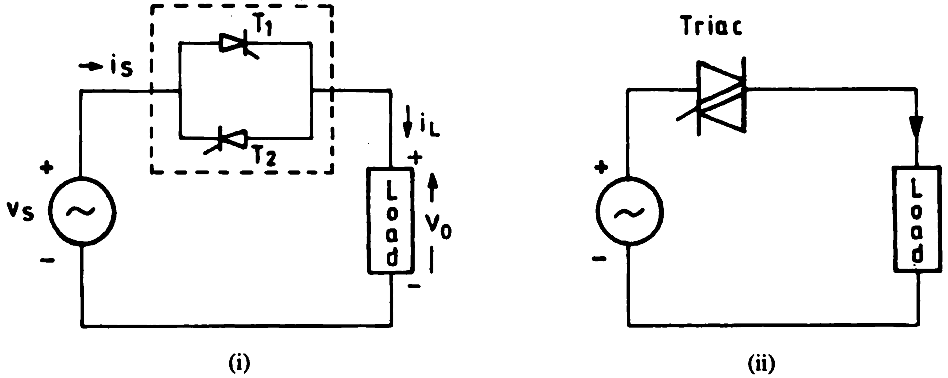
4.6 CYCLOCONVERTER

A cycloconverter can be used mainly for converting power at low frequency from line frequency supply. This is suitable for low speed drives, including synchronous motors.

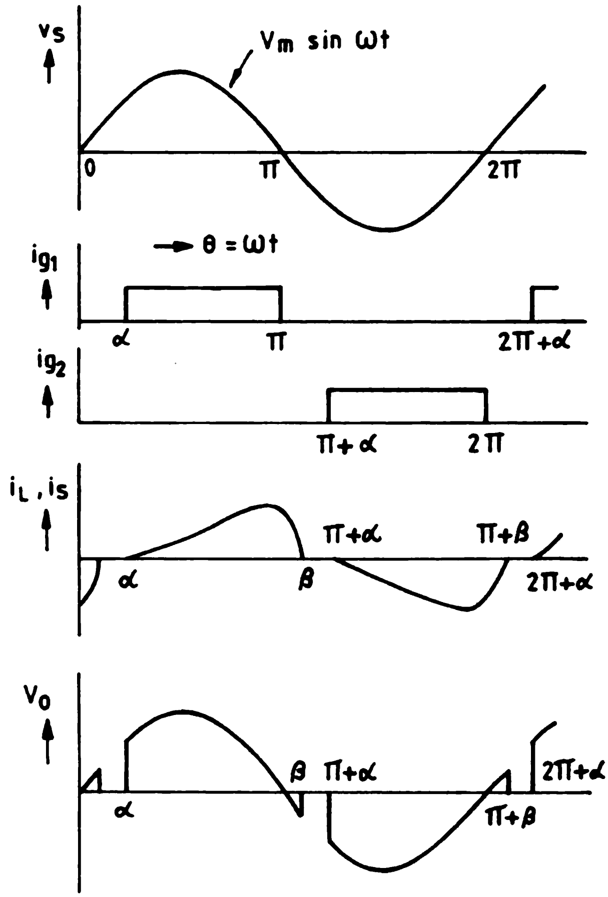
A three-phase to single-phase cycloconverter, along with the output waveform, is shown in Fig. 4.21. Two three-phase half-wave circuits are connected in parallel, as in a dual converter scheme shown in Fig. 3.14. One group termed positive conducts, when the polarity of output voltage in positive, while the other, the negative group conducts for the negative half of the output voltage. In Fig. 4.21c, the firing delay angle α at the point A is 90°, and the firing angle is gradually reduced till it reaches the point G, where the output voltage increases to maximum. In between the points B to G, the firing angle is $90° > \alpha > 0°$. The firing angle is again increased to 90°, when the voltage is reduced from maximum value to zero. The negative half of the waveform is generated, when the other group (N-group) is conducting. The firing angle in this case varies from 90° to 180° and again decreases to 90°. If the load power factor is lagging, the load current lags behind the voltage. For the case, when the current is negative and voltage positive, the current flows through the N-group and also for the next half when the current is positive and voltage negative, thyristors in the P-group conduct. A three-phase to three-phase cycloconverter has six bridge circuits, each having six thyristors. Two thyristor bridge circuits per phase are needed.

Cycloconverter-fed drive

Cycloconverters are generally used for low speed, high power synchronous motor drives. The output voltage at low frequency contains less harmonics, thus resulting in smooth operation. The harmonic



(a) Power circuits.



(b) Voltage and current waveforms.

Fig. 4.20 Single-phase ac voltage controller with inductive (R-L) load ($\alpha > \phi$).

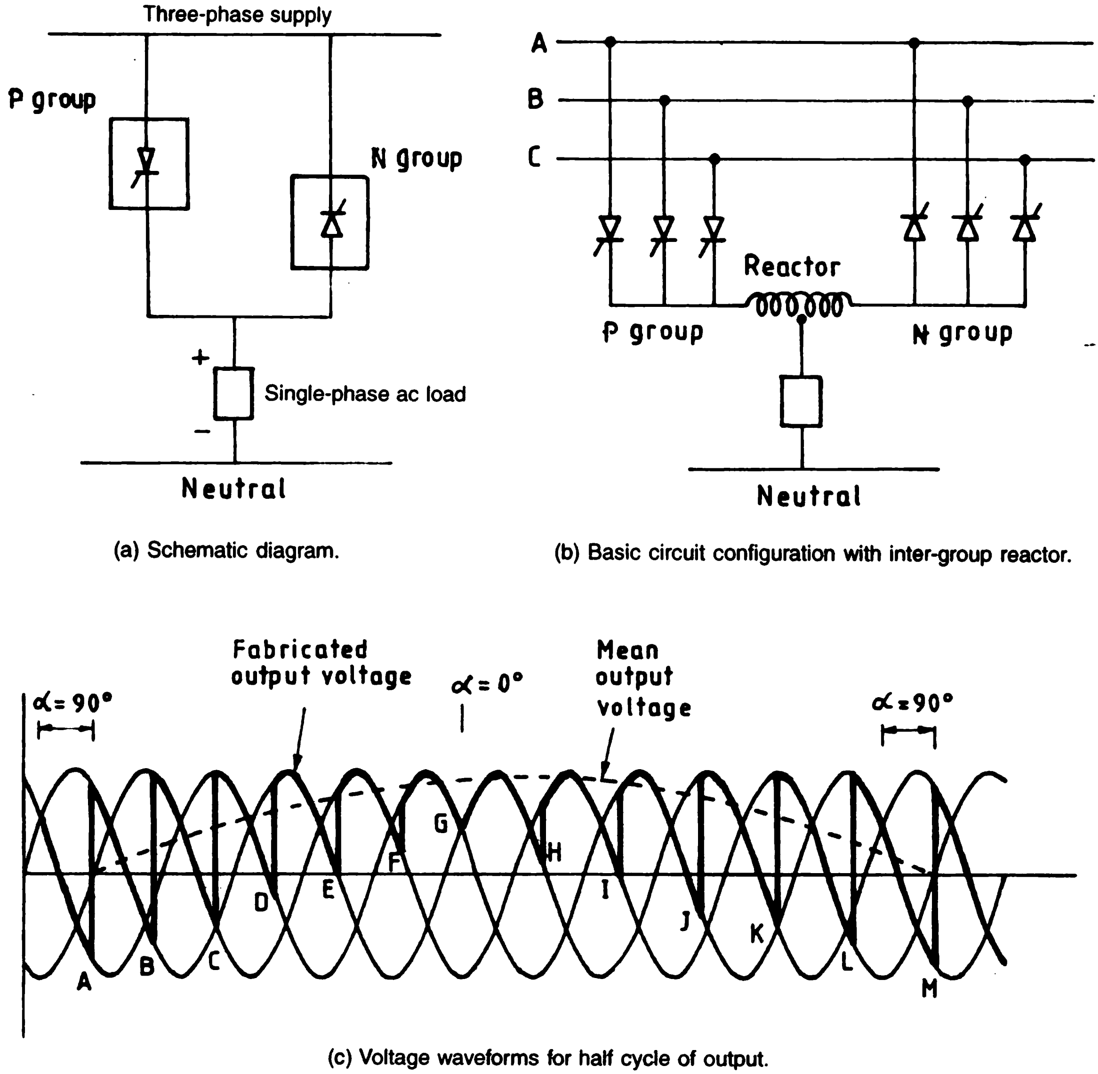


Fig. 4.21 Three-phase to single-phase cycloconverter.

content increases, resulting in jerky motion, as the frequency is increased, which is normally limited to nearly 1/3 of input frequency. Also, as a large number of thyristors are required, cycloconverters are cost effective only for large power drives.

4.7 SPEED CONTROL OF SYNCHRONOUS MOTORS

4.7.1 Input Frequency Variation

As in the case of Induction motors, synchronous motors are fed from a variable frequency source such as a voltage source inverter as discussed earlier in respect of speed control. The control strategy is general for ac motors, whether induction or synchronous, as the voltage to frequency ratio is proportional to flux. Therefore, a constant voltage to frequency (V/f) ratio is maintained for frequencies below base (or rated) frequency, and the ratio is increased at low speeds to take into account the stator resistance drop. This may be termed variable-voltage variable-frequency (VVVF) control. This also results in a constant pull-out torque at all frequencies. Similarly, for frequencies above base frequency, a constant voltage is applied, which results in decrease of flux and also developed torque; otherwise, a voltage higher than rated value is needed to keep flux constant, which may not be permitted in most cases. For this case, the motor may be operated in two modes: (a) True synchronous mode, and (b) Self-controlled mode.

In true synchronous mode, the motor is supplied from a variable frequency inverter (voltage source) with constant V/f ratio. The frequency is slowly increased such that the difference between the synchronous speed and rotor speed is small and the machine speed increases at a slow rate to keep pace with the change in input frequency or synchronous speed. This type of control is used both for speed control and starting as well as braking, the only limitation as stated is the slow rate of change in frequency. Braking occurs when the rotor speed is more than the synchronous speed; the energy is fed back to ac supply from the motor, and the inverter or the supply, therefore, must be capable of regeneration.

Such drives using reluctance or permanent magnet types of motors can be used for multi-machine drives in textile and paper mills, fed by one voltage source inverter only, where accurate tracking of motor speeds is required.

In the self-controlled mode, the stator supply frequency is changed with the rotor speed, which remains always at synchronous speed, as the rotor position is sensed and accordingly, the switching sequence of devices in the inverter, used for the stator, is determined by the control circuit. So, the rotor is always in step with input frequency and cannot pull out as in the earlier case. The motor does not require a damper winding. This is termed dc 'brushless' motor and the input supply is dc voltage in the case of inverter. The input voltage (dc) is varied to control the speed. Other types of ac 'brushless' motors can be implemented using a cycloconverter, which produces low output frequency for motors required to be run at low speed. Thus, cycloconverters are suitable for high power, low speed drives used for cement kilns and mine-hoists, including other special applications. The firing pulses are obtained by rotor position sensors (optical or Hall effect type) or by comparison of motor terminal voltages.

4.7.2 Self-controlled Synchronous Motor Drive using Load Commutated Thyristor Inverter

A self-controlled synchronous motor (SCSM) drive using a load commutated thyristor inverter is shown in Fig. 4.22. Two three-phase bridge rectifiers (converters) are needed. One of them connected on the source side, acts as a line-commutated converter with $0 < \alpha < \pi/2$ to obtain dc output voltage from ac input supply, while the other connected on the load side, acts as a load-commutated inverter with $\pi/2 < \alpha < \pi$ to obtain three-phase ac output to be fed to the motor. In the latter case, the thyristors are commutated by the application of motor induced voltages, with input power factor taken as leading. The delay angle α is measured by comparison of induced voltages in the same way as in the case of line voltage in a line-commutated converter. This is applicable for motoring operation. For braking operation, the converter on the load side is used to obtain dc voltage as in the line-commutated case, while the converter on the source side acts as a line-commutated inverter

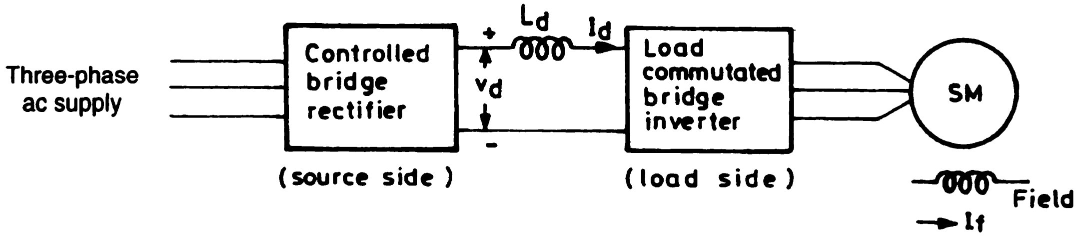


Fig. 4.22 Self-controlled synchronous motor drive using load commutated thyristor inverter.

to feed power to ac supply. The delay angle range changes accordingly as the operation in one case changes from acting as converter to inverter, while in the other case the change is from inverter to converter operation. The firing pulses are obtained either by comparison of motor terminal voltages (as induced voltages are not directly accessible) or by rotor position sensors as stated earlier. The advantages of load commutation over forced commutation are: (i) the absence of commutation circuits needed for thyristors, (ii) operation at higher frequency, and (iii) operation at higher power levels. Certain problems are encountered for operation at low speeds which can be overcome by operating the source side converter as inverter in pulsed mode, as the load side inverter frequency (and the voltage induced) is low compared to high frequency in high speed operation.

Example 4.1

The rotor of an eight-pole, 50 Hz, three-phase induction motor has a resistance of 0.2 Ω per phase and runs at 720 rpm. If the load torque remains unchanged, calculate the additional rotor resistance that will reduce its speed by 10%. Neglect stator impedance.

Solution

$f = 50$ Hz, $p = 8$, $N_1 = 720$ rpm, $r_2 = 0.2\ \Omega$

$$N_s = \frac{120f}{p} = \frac{120 \times 50}{8} = 750 \text{ rpm}$$

$$s_1 = \frac{N_s - N_1}{N_s} = \frac{750 - 720}{750} = 0.04$$

The speed is reduced by 10% to

$$N_2 = (1 - 0.1) \times 720 = 648 \text{ rpm}$$

$$s_2 = \frac{N_s - N_2}{N_s} = \frac{750 - 648}{750} = 0.136$$

Let the total resistance including additional rotor resistance inserted in the rotor circuit be

$$R_2 = r_2 + R_e$$

The torque is constant $\left(T \propto \frac{s_1}{r_2}\right)$, therefore

$$\frac{s_1}{r_2} = \frac{s_2}{R_2}$$

or

$$R_2 = \frac{s_2 r_2}{s_1} = \frac{0.136 \times 0.2}{0.04} = 0.68\ \Omega$$

Hence, the resistance to be inserted in the rotor circuit is

$$R_e = R_2 - r_2 = 0.68 - 0.2 = 0.48\ \Omega$$

Example 4.2

The rotor of a six-pole, 50 Hz, slip-ring induction motor has a resistance of 0.2 Ω per phase and runs at 960 rpm on full load. Calculate the resistance per phase to be inserted in the rotor circuit such that the speed is reduced to 800 rpm, the torque being (a) the same as before, and (b) proportional to square of the speed.

Solution

$f = 50$ Hz, $p = 6$, $r_2 = 0.2\ \Omega$, $N_1 = 960$ rpm, $N_2 = 800$ rpm

$$N_s = \frac{120f}{p} = \frac{120 \times 50}{6} = 1000 \text{ rpm}$$

$$s_1 = \frac{N_s - N_1}{N_s} = \frac{1000 - 960}{1000} = 0.04$$

Neglecting $(s_1 x_2)$ compared to r_2, we get $I_2 = \frac{s_1 E_2}{r_2}$.

Full load electromagnetic torque

$$T_{L1} = \frac{3I_2^2 r_2}{\omega_s s_1} = \frac{3E_2^2 s_1}{\omega_s r_2}$$

$$s_2 = \frac{N_s - N_2}{N_s} = \frac{1000 - 800}{1000} = 0.2$$

Let the total resistance including additional rotor resistance inserted in the rotor circuit be

$$R_2 = r_2 + R_e$$

Then

$$T_{L2} = \frac{3E_2^2 s_2}{\omega_s R_2}$$

(a) $T_{L2} = T_{L1}$ (T_L = constant). Therefore

$$\frac{s_1}{r_2} = \frac{s_2}{R_2}$$

or

$$R_2 = \frac{s_2 r_2}{s_1} = \frac{0.2 \times 0.2}{0.04} = 1.0\ \Omega$$

Hence, the additional resistance to be inserted in the rotor circuit is

$$R_e = R_2 - r_2 = 1.0 - 0.2 = 0.8\ \Omega$$

(b) $T_L \propto N^2$. Therefore

$$\frac{T_{L2}}{T_{L1}} = \left(\frac{N_2}{N_1}\right)^2$$

or

$$\frac{s_2 r_2}{R_2 s_1} = \left(\frac{N_2}{N_1}\right)^2$$

or

$$R_2 = \frac{s_2 r_2}{s_1}\left(\frac{N_1}{N_2}\right)^2 = \frac{0.2 \times 0.2}{0.04} \times \left(\frac{960}{800}\right)^2 = 1.44\ \Omega$$

Therefore

$$R_e = R_2 - r_2 = 1.44 - 0.2 = 1.24\ \Omega$$

Example 4.3

A 220 V, three-phase, four-pole, 60 Hz, squirrel cage induction motor develops a maximum internal torque of 250% at a slip of 16%, when operating at rated voltage and frequency. If the effect of stator impedance is neglected, determine the maximum internal torque that the motor would develop, if it were operated at 200 V, 50 Hz. Under this condition, at what speed in rpm, would the maximum torque be developed? Find also the full-load speed for 220 V, 60 Hz supply.

Solution

$$V_1 = 220\text{ V}, \quad f_1 = 60\text{ Hz}, \quad p = 4, \quad s_{m1} = \frac{r_2'}{x_2'} = 0.16$$

$$V_2 = 200\text{ V}, \quad f_2 = 50\text{ Hz}, \quad \frac{V_2}{V_1} = \frac{200}{220} = 0.9091$$

Rotor reactance referred to the stator at 60 Hz (f_1) = x_2'.

Rotor reactance referred to the stator at 50 Hz (f_2) = $x_2'' = \frac{f_2}{f_1}x_2' = \frac{50}{60}x_2' = 0.833x_2'$

$$s_{m2} = \frac{r_2'}{x_2''} = \frac{r_2'}{x_2'}\cdot\frac{x_2'}{x_2''} = \frac{0.16}{0.833} = 0.192$$

$$T_{m1} = \frac{3V_1^2}{2x_2'} = 2.5T_{fl}$$

$$T_{m2} = \frac{3V_2^2}{2x_2''} = \frac{3V_1^2}{2x_2'}\left(\frac{V_2}{V_1}\right)^2\cdot\left(\frac{x_2'}{x_2''}\right) = \frac{(0.9091)^2}{0.833}T_{m1} = (0.992 \times 2.5)T_{fl} = 2.48T_{fl}$$

$$N_{s1} = \frac{120f_1}{p} = \frac{120}{4}f_1 = 30f_1 = 30 \times 60 = 1800 \text{ rpm}$$

Speed at maximum torque $N_{m1} = (1 - s_{m1}) \cdot N_{s1} = N_{s1} - N_{sl1}$

$$= (1 - 0.16)1800 = 1800 - 288 = 1512 \text{ rpm}$$

$$N_{s2} = \frac{120f_2}{p} = 30f_2 = 30 \times 50 = 1500 \text{ rpm}$$

Speed at maximum torque $N_{m2} = (1 - s_{m2})\, N_{s2} = N_{s2} - N_{sl2}$

$$= (1 - 0.192)1500 = 1500 - 288 = 1212 \text{ rpm}$$

Now

$$T_{fl} = \frac{2T_{m1}}{\dfrac{s_1}{s_{m1}} + \dfrac{s_{m1}}{s_1}} = \frac{2 \times 2.5 \times T_{fl}}{\dfrac{s_1}{0.16} + \dfrac{0.16}{s_1}}$$

or

$$\frac{s_1}{0.16} + \frac{0.16}{s_1} = 5.0$$

Solving the above equation, we get $s_1 = 0.0334$. Therefore

$$N_1 = (1 - s_1)N_{s1} = (1 - 0.0334)1800 = 1740 \text{ rpm}$$

Example 4.4

A three-phase bridge inverter is used to run a three-phase induction motor rated at 440 V, 15 A, 1440 rpm. The maximum to minimum speed ratio required is 10:1. Find the minimum and maximum dc input voltage for the inverter. The inverter is operated in 180° conduction mode. If this voltage is to be obtained from a three-phase full-controlled bridge converter fed from 440 V mains, calculate the firing angles as needed.

Solution

$N_r = 1440$ rpm, $f_1 = 50$ Hz, $N_r \approx N_s$

$$p = \frac{120 f_1}{N_s} \approx \frac{120 f_1}{N_r} = \frac{120 \times 50}{1440} = 4.167 \approx 4 \text{ (even no.)}$$

For the three-phase bridge inverter in 180° conduction mode, we have

$$V_{ac} = \frac{2\sqrt{2}}{\pi} V_{dc} \cos\left(\frac{\pi}{6}\right)$$

or

$$V_{dc} = \frac{440}{(0.9 \times 0.866)} = 564.5 \text{ V (maximum), for } f_1 = 50 \text{ Hz}$$

The frequency of the inverter is to be 5–50 Hz for the speed control required.

To maintain constant flux in the air gap of the motor, the motor input (inverter output) voltage at the lowest frequency $f_2 = 5$ Hz is

$$V'_{ac} = \frac{f_2}{f_1} V_{ac} = \frac{5}{50} \times 440 = 0.1 \times 440 = 44 \text{ V}$$

The input voltage to the inverter is

$$V'_{dc} = 0.1 \times 564.5 = 56.45 \text{ V (minimum)}$$

For the three-phase full-controlled bridge converter, we have

$$V_{dc} = \frac{3\sqrt{2}}{\pi} V_{ac} \cos \alpha_1 = (1.35 \times 440) \cos \alpha_1 = 594 \times \cos \alpha_1$$

or

$$\alpha_1 = \cos^{-1}\left(\frac{564.5}{594.0}\right) = 18.25°, \text{ for maximum } V_{dc}$$

$$\alpha_2 = \cos^{-1}\left(\frac{56.45}{594.0}\right) = 84.55°, \text{ for minimum } V_{dc}$$

Example 4.5

Consequent pole method is used to obtain the following synchronous speeds at 50 Hz:

1000 rpm, 750 rpm, 500 rpm and 375 rpm.

Find the number of standard poles and consequent poles for each connection.

Solution

By consequent pole technique, the number of poles may be increased in the ratio 1:2, or in other words the speed may be changed in the ratio 2:1.

1000 rpm, and 500 rpm for one winding, $f = 50$ Hz

$$p = \frac{120f}{N}, \text{ where } p \text{ is the no. of poles}$$

$$\text{Standard poles, } p_1 = \frac{120 \times 50}{1000} = 6$$

$$\text{Consequent poles, } p_2 = \frac{120 \times 50}{500} = 12$$

750 rpm, and 375 rpm for the other winding, $f = 50$ Hz

$$\text{Standard poles, } p_3 = \frac{120 \times 50}{750} = 8$$

$$\text{Consequent poles, } p_4 = \frac{120 \times 50}{375} = 16$$

Example 4.6

A three-phase, 400 V, 50 Hz, six-pole, 925 rpm, star-connected, induction motor has the following parameters:

$$r_1 = 0.2\ \Omega, \quad r_2' = 0.3\ \Omega, \quad x_1 = 0.5\ \Omega, \quad x_2' = 1.0\ \Omega$$

The motor is fed from a voltage source inverter with a constant voltage to frequency, *V/f* ratio.

(a) Calculate the maximum torque T_m and the corresponding speed N_m for 50 Hz and 20 Hz. Do the above values change, if r_1 is neglected?

(b) Find the *V/f* ratio at 20 Hz, so that the maximum torque at this frequency remains the same as at 50 Hz.

Solution

$$r_1 = 0.2\ \Omega, \quad r_2' = 0.3\ \Omega, \quad x_1 = 0.5\ \Omega, \quad x_2' = 1.0\ \Omega$$

$$x_1 + x_2' = 0.5 + 1.0 = 1.5\ \Omega, \qquad \text{pole-pairs, } p = \frac{6}{2} = 3$$

(a) For $f_1 = 50$ Hz, $\quad V_{s1} = \dfrac{400}{\sqrt{3}} = 231$ V

$$\omega_{s1} = \frac{2\pi f_1}{p} = \frac{2\pi \times 50}{3} = 104.72 \text{ rad/s}, \qquad N_{s1} = \frac{60}{2\pi} 104.72 = 1000 \text{ rpm}$$

$$s_{m1} = \frac{r_2'}{\sqrt{(r_1)^2 + (x_1 + x_2')^2}} = \frac{0.3}{\sqrt{(0.2)^2 + (1.5)^2}} = \frac{0.3}{1.513} = 0.198$$

$$T_{m1} = \frac{3V_{s1}^2}{2\omega_{s1}\left(r_1 + \sqrt{(r_1)^2 + (x_1 + x_2')^2}\right)} = \frac{3 \times (231)^2}{2 \times 104.72 \times (0.2 + 1.513)} = 446 \text{ N} \cdot \text{m}$$

$$N_{m1} = (1 - s_{m1})N_{s1} = (1 - 0.198) \times 1000 = 802.0 \text{ rpm}$$

For $f_2 = 20$ Hz, $\alpha = \frac{f_2}{f_1} = \frac{20}{50} = 0.4$

$$s_{m2} = \frac{\left(\frac{r_2'}{\alpha}\right)}{\sqrt{\left(\frac{r_1}{\alpha}\right)^2 + (x_1 + x_2')^2}} = \frac{\left(\frac{0.3}{0.4}\right)}{\sqrt{\left(\frac{0.2}{0.4}\right)^2 + (1.5)^2}} = \frac{0.75}{1.581} = 0.4743$$

$$T_{m2} = \frac{3V_{s1}^2}{2\omega_{s1}\left[\left(\frac{r_1}{\alpha}\right) + \sqrt{\left(\frac{r_1}{\alpha}\right)^2 + (x_1 + x_2')^2}\right]} = \frac{3 \times (231)^2}{2 \times 104.72 \times \left(\frac{0.2}{0.4} + 1.581\right)} = 367.3 \text{ N} \cdot \text{m}$$

$$N_{m2} = (1 - s_{m2})N_{s2} = (1 - 0.4743) \times 0.4 \times 1000 = 210 \text{ rpm}$$

If r_1 is neglected, $f_1 = 50$ Hz,

$$s'_{m1} = \frac{r_2'}{(x_1 + x_2')} = \frac{0.3}{1.5} = 0.2$$

$$T'_{m1} = \frac{3V_{s1}^2}{2\omega_{s1}(x_1 + x_2')} = \frac{3 \times (231)^2}{2 \times 104.72 \times 1.5} = 509.56 \text{ N} \cdot \text{m}$$

$$N'_{m1} = (1 - s'_{m1})N_{s1} = (1 - 0.2) \times 1000 = 800 \text{ rpm}$$

For $f_2 = 20$ Hz

V/f is constant, as input voltage V_{s2} is reduced to ($0.4\ V_{s1}$), N_{s2} also reduces to ($0.4\ N_{s1}$).

s'_{m2} remains the same as s'_{m1}, being independent of frequency.

$N'_{m2} = (1 - s'_{m2})N_{s2} = (1 - 0.2) \times 0.4 \times 1000 = 320$ rpm

T'_{m2} remains the same as T'_{m1}, also being independent of frequency.

(b) Let us take input voltage, V''_{s2} as $(\beta \cdot V_{s1})$ so as to keep T''_{m2} same as T_{m1}.

$$T''_{m2} = T_{m1} = \frac{3(V_{s1})^2\left(\frac{\beta}{\alpha}\right)^2}{2\omega_{s1}\left[\left(\frac{r_1}{\alpha}\right)+\sqrt{\left(\frac{r_1}{\alpha}\right)^2+(x_1+x_2')^2}\right]} = \left(\frac{\beta}{\alpha}\right)^2 367.3 = 446$$

or

$$\frac{\beta}{\alpha} = \sqrt{\frac{446}{367.3}} = 1.102$$

or

$$\beta = 1.102\alpha = 1.102 \times 0.4 = 0.441$$

Now

$$\frac{V_{s1}}{f_1} = \frac{231}{50} = 4.62$$

and

$$\frac{V''_{s2}}{f_2} = \frac{\beta \cdot V_{s1}}{f_2} = \frac{0.441 \times 231}{20} = 5.093$$

Therefore, the new V/f ratio required is 5.092.

Note. (i) Neglecting r_1 may introduce a significant error in torque computation, especially at low frequency.

(ii) The V/f ratio is to be increased to maintain maximum torque constant, as frequency is decreased.

Example 4.7

The motor of Example 4.6 is to be braked by plugging from its initial full-load speed of 925 rpm. The stator to rotor turns ratio is 2. Calculate:

(a) the initial braking current and torque as a ratio of their full-load values,

(b) the value of resistance required to be inserted in rotor circuit to reduce the maximum braking current to 1.5 times full-load current and also the initial braking torque, and

(c) the value of resistance required to be inserted in rotor circuit and also the maximum braking current, when the maximum braking torque is made to occur at 600 rpm.

Solution

$N_s = 1000$ rpm, $\omega_s = 104.72$ rad/s, $N_m = 925$ rpm, $r_1 = 0.2\ \Omega$,

$r_2' = 0.3\ \Omega$, $x_1 = 0.5\ \Omega$, $x_2' = 1.0\ \Omega$

$$x_1 + x_2' = 0.5 + 1.0 = 1.5\ \Omega, \qquad V_s = \frac{400}{\sqrt{3}} = 231\ \text{V}$$

$$s_1 = \frac{N_s - N_m}{N_s} = \frac{1000 - 925}{1000} = 0.075, \qquad \frac{r_2'}{s_1} = \frac{0.3}{0.075} = 4.0$$

$$I_1 = \frac{V_s}{\sqrt{\left(r_1 + \frac{r_2'}{s_1}\right)^2 + (x_1 + x_2')^2}} = \frac{231}{\sqrt{(0.2 + 4.0)^2 + (1.5)^2}} = 51.8\ \text{A}$$

$$T_1 = \frac{3(I_1)^2\left(\frac{r_2'}{s_1}\right)}{\omega_s} = \frac{3 \times (51.8)^2 \times 4.0}{104.72} = 307.5\ \text{N}\cdot\text{m}$$

(a) Plugging

$$s_{pl} = 2 - s_1 = 2 - 0.075 = 1.925, \qquad \frac{r_2'}{s_{pl}} = \frac{0.3}{1.925} = 0.156$$

Initial braking (plugging) current, $I_{pl} = \dfrac{231}{\sqrt{(0.2 + 0.156)^2 + (1.5)^2}} = 149.84$ A

Therefore

$$\frac{I_{pl}}{I_1} = \frac{149.84}{51.8} = 2.89$$

Initial braking torque, $T_{pl} = \dfrac{3 \times (149.84)^2 \times 0.156}{104.72} = 100.34$ N · m

Therefore

$$\frac{T_{pl}}{T_1} = \frac{100.34}{307.5} = 0.326$$

It may be noted that, while current has increased by nearly three times, the torque has reduced by a factor of 0.326.

(b) With an external resistance R_e in the rotor circuit, and $I_{pl}' = 1.5\, I_1 = 1.5 \times 51.8 = 77.7$ A, we have

$$I_{pl}' = \frac{V_s}{\sqrt{\left(r_1 + \frac{(r_2' + R_e')}{S_{pl}}\right)^2 + (x_1 + x_2')^2}} = \frac{231.0}{\sqrt{\left(0.2 + \frac{(0.3 + R_e')}{1.925}\right)^2 + (1.5)^2}} = 77.7\ \text{A}$$

or

$$\left(0.356+\frac{R'_e}{1.925}\right)^2+(1.5)^2=\left(\frac{231.0}{77.7}\right)^2=8.84$$

or

$$R'_e = 1.925 \times (2.567 - 0.356) = 4.256\ \Omega$$

Therefore

External resistance, $R_e = \dfrac{R'_e}{a^2} = \dfrac{4.256}{(2)^2} = 1.064\ \Omega$ ($\because$ stator to rotor turns ratio, $a = 2$)

and

$$T'_{pl} = \frac{3(I'_{pl})^2\left(\dfrac{r'_2+R'_e}{s_{pl}}\right)}{\omega_s} = \frac{3\times(77.7)^2\times\left(\dfrac{0.3+4.256}{1.925}\right)}{104.72} = 409.34\ \text{N}\cdot\text{m}$$

Therefore

$$\frac{T'_{pl}}{T_1} = \frac{409.34}{307.4} = 1.33$$

The use of an external resistance in the rotor circuit reduces current from nearly 3 to 1.5 times full-load current; the torque is increased from 0.326 to 1.33 times full-load torque.

(c) $N_2 = -600$ rpm, $\quad s'_{pl} = \dfrac{N_s - N_2}{N_s} = \dfrac{1000-(-600)}{1000} = 1.6$

As the maximum braking torque is at speed N_2, we have

$$\frac{r'_2 + R'_e}{s'_{pl}} = \sqrt{(r_1)^2 + (x_1 + x'_2)^2}$$

or

$$\frac{0.3 + R'_e}{1.6} = \sqrt{(0.2)^2 + (1.5)^2} = 1.513$$

Therefore

$$R'_e = (1.6 \times 1.513) - 0.3 - 2.12\ \Omega$$

or

$$R_e = \frac{R'_e}{a^2} = \frac{2.12}{(2)^2} = 0.53\ \Omega$$

$$s_{pl} = 1.925, \qquad \therefore \qquad \frac{r'_2 + R'_e}{s_{pl}} = \frac{0.3 + 2.12}{1.925} = 1.257\ \Omega$$

Maximum braking current with external resistance R_e (= 0.53 Ω) inserted in rotor circuit,

$$I''_{pl} = \frac{V_s}{\sqrt{\left(r_1 + \dfrac{r'_2 + R'_e}{s_{pl}}\right)^2 + (x_1 + x'_2)^2}} = \frac{231}{\sqrt{(0.2 + 1.257)^2 + (1.5)^2}} = 110.5\ \text{A}$$

Example 4.8

A three-phase, 400 V, 50 Hz, six-pole, 960 rpm, star-connected wound rotor induction motor has the following constants referred to the stator:

$$r_1 = 0.5\ \Omega, \qquad r_2' = 0.7\ \Omega, \qquad x_1 = x_2' = 1.5\ \Omega$$

The motor drives a fan load at 960 rpm. The stator to rotor turns ratio is 2. Calculate the resistance required to be connected in each phase of the rotor circuit to reduce the speed to 600 rpm.

Solution

$$r_1 = 0.5\ \Omega, \qquad r_2' = 0.7\ \Omega, \qquad x_1 = x_2' = 1.5\ \Omega, \qquad x_1 + x_2' = 1.5 + 1.5 = 3.0\ \Omega$$

$$\text{pole-pairs, } p = \frac{6}{2} = 3, \qquad V_s = \frac{400}{\sqrt{3}} = 231\text{ V}, \qquad f = 50\text{ Hz}, \qquad N_1 = 960\text{ rpm}$$

$$\omega_s = \frac{2\pi f}{p} = \frac{2\pi \times 50}{3} = 104.72\text{ rad/s}, \qquad N_s = \frac{60}{2\pi}104.72 = 1000\text{ rpm}$$

$$s_1 = \frac{N_s - N_1}{N_s} = \frac{1000 - 960}{1000} = 0.04, \qquad \frac{r_2'}{s_1} = \frac{0.7}{0.04} = 17.5$$

$$I_1 = \frac{V_s}{\sqrt{\left(r_1 + \frac{r_2'}{s_1}\right)^2 + (x_1 + x_2')^2}} = \frac{231.0}{\sqrt{(0.5 + 17.5)^2 + (3.0)^2}} = 12.66\text{ A}$$

$$T_1 = \frac{3(I_1)^2\left(\frac{r_2'}{s_1}\right)}{\omega_s} = \frac{3 \times (12.66)^2 \times 17.5}{104.72} = 80.35\text{ N}\cdot\text{m},$$

$$T \propto N^2, \text{ for fan load,} \qquad N_2 = 600\text{ rpm}$$

$$T_2 = \left(\frac{N_2}{N_1}\right)^2 T_1 = \left(\frac{600}{960}\right)^2 \times 80.35 = 31.38\text{ N}\cdot\text{m}$$

$$s_2 = \frac{N_s - N_2}{N_s} = \frac{1000 - 600}{1000} = 0.4$$

As external resistance R_e per phase is inserted in the rotor circuit, the term,

$\frac{r_2'}{s_1}$ is replaced by $y = \frac{(r_2' + R_e')}{s_2} = \frac{(0.7 + R_e')}{0.4}$ in the expressions for I_1 and T_1.

Therefore

$$T_2 = \frac{3V_s^2 y}{\omega_s[(r_1 + y)^2 + (x_1 + x_2')^2]}$$

or

$$T_2 = \frac{3 \times (231)^2 \times y}{104.72 \times (0.5 + y)^2 + (3)^2} = 31.38$$

or

$$\frac{y}{(0.5 + y)^2 + 9} = \frac{31.38 \times 104.72}{3 \times (231)^2} = \frac{1093.77}{(231)^2} = 0.02053$$

or

$$y^2 - 47.712y + 9.25 = 0$$

or

$$y = 47.517 \quad \text{or} \quad 0.195$$

Therefore

$$y = \frac{0.7 + R_e'}{0.4} = 47.517$$ (neglecting the smaller value of y as it gives negative value of R_e)

or

$$R_e' = (47.517 \times 0.4) - 0.7 = 18.3\ \Omega$$

Therefore

$$R_e = \frac{R_e'}{a^2} = \frac{18.3}{(2)^2} = 4.575\ \Omega$$ ($\because$ stator to rotor turns ratio $a = 2$)

Example 4.9

A 400 kW, three-phase, 3.3 kV, 50 Hz, unity power factor, four-pole, star-connected synchronous motor has the following parameters:

$$r_a = 0, \qquad x_s = 12\ \Omega, \qquad \text{rated field current} = 10\ \text{A}$$

The machine is controlled by variable frequency control at a constant V/f ratio. Calculate:

(a) the torque and field current for rated armature current, 900 rpm and 0.8 leading power factor, and

(b) the armature current and power factor for regenerative braking torque equal to rated motor torque, 900 rpm and rated field current.

Solution

$r_a = 0$, $x_s = 12\ \Omega$, $p = 4$, $f = 50$ Hz, $P = 400$ kW

Rated field current, $I_f = 10$ A, $V_s = \frac{3300}{\sqrt{3}} = 1905$ V

Rated armature current, $I_s = \dfrac{P}{3V_s} = \dfrac{400 \times 10^3}{3 \times 1905} = 70.0$ A

(a) $E = V_s - jx_sI_s = 1905 - 12\ \angle 90° \times 70\angle 0° = 1905 - j840 = 2082\ \text{V}\ \angle -23.8°$

Synchronous speed, $N = \dfrac{120f}{p} = \dfrac{120 \times 50}{4} = 1500$ rpm

For operation at speed N' of 900 rpm, we have

$$f' = \frac{N'}{N}f = \frac{900}{1500} \times 50 = 0.6 \times 50 = 30 \text{ Hz}$$

For 0.8 power factor leading, $\cos \phi' = 0.8$, $\phi' = 36.87°$

$$x_s' = \frac{f'}{f}x_s = \frac{30}{50} \times 12 = 7.2\ \Omega$$

$$V_s' = \frac{f'}{f}V_s = 0.6 \times 1905 = 1143 \text{ V}$$

$E' = V_s' - jx_s'I_s' = 1143 - 7.2\angle 90° \times 70\ \angle 36.87° = 1445.4 - j403.2 = 1501\ \text{V}\ \angle -15.6°$

At rated field current I_f, and speed N', $E'' = \dfrac{f'}{f}E = 0.6 \times 2082 = 1249.2$ V

Field current, $I_f' = \dfrac{E'}{E''}I_f = \dfrac{1501.0}{1249.2} \times 10 = 12.0$ A

Power input $P' = 3V_s'I_s' \cos \phi' = 3 \times 1143 \times 70 \times 0.8 = 192$ kW

Motor speed, $\omega' = \dfrac{2\pi}{60}N' = \dfrac{2\pi}{60} \times 900 = 94.25$ rad/s

Torque, $T' = \dfrac{P'}{\omega'} = \dfrac{192 \times 10^3}{94.25} = 2037.1\ \text{N} \cdot \text{m}$

(b) At speed N', rated field current I_f, and also rated torque, we have

Output power, $P'' = \dfrac{N'}{N}P = 0.6 \times 400 = 240\ \text{kW}$

$$V_s' = 1143 \text{ V} \qquad x_s' = 7.2\ \Omega \qquad E'' = 1249.2 \text{ V}$$

$$P'' = \frac{3V_s'E''}{x_s'}\sin \delta$$

or

$$240 \times 10^3 = \frac{3 \times 1143 \times 1249.2}{7.2} \sin \delta = (595.1 \times 10^3) \sin \delta$$

or

$$\sin \delta = \frac{240.0 \times 10^3}{595.1 \times 10^3} = 0.4033$$

Therefore

$$\delta = 23.8°$$

Also

$$I_s'' = \frac{E'' - V_s'}{jx_s'} = \frac{1249.2 \angle 23.8° - 1143 \angle 0°}{7.2 \angle 90°} = 70\ A \angle 0°$$

or

$$I_s'' = 70.0\ \text{A}$$

and

$$\text{power factor} = \cos 0° = 1.0 \text{ (unity)}$$

Example 4.10

Two induction motors are connected in cascade electrically and mechanically coupled by a gear train so that the speed of the first machine is 'a' times the speed of the second (Fig. 4.23). Show that the synchronous speed of the combination is $n = af/(ap_1 + p_2)$, where f is the line frequency, and p_1 and p_2 are the number of pole-pairs of the two machines respectively.

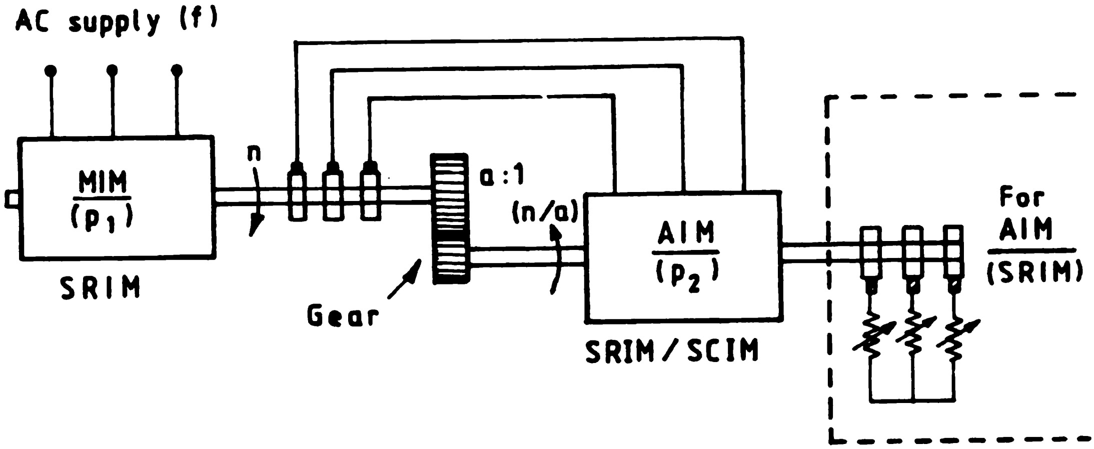


Fig. 4.23 Cascade connection of two induction motors (Example 4.10).

Solution

The synchronous speed of the first motor is $n_1 = \dfrac{f}{p_1}$.

Let the speed of the first machine be n. The rotor frequency of the first machine and the stator frequency of the second machine are,

$$f_2 = p_1(n_1 - n)$$

The synchronous speed of the second machine is

$$n_2 = \frac{f_2}{p_2} = \frac{p_1}{p_2}(n_1 - n)$$

The rotor speed of the second machine is given as $\dfrac{n}{a}$. The rotor frequency of the second machine is

$$p_2\left(n_2 - \frac{n}{a}\right) \cong 0$$

or

$$n_2 \cong \frac{n}{a}$$

Hence

$$\frac{n}{a} = \frac{p_1}{p_2}(n_1 - n) = \frac{p_1}{p_2}\left(\frac{f}{p_1} - n\right)$$

or

$$n = \frac{af}{(ap_1 + p_2)} = \frac{f}{\left(p_1 + \dfrac{p_2}{a}\right)}$$

Example 4.11

An induction motor, whose windings can be arranged for 12 or 24 poles, is run in cascade with a direct coupled induction motor, which can be arranged for 2 or 4 poles. Calculate the approximate speeds, which can be obtained from the set, the supply frequency being 50 Hz.

Solution

Main motor. 12 or 24 poles, i.e. 6 or 12 pole-pairs (p_1), output power, P_1 (W).
Auxiliary motor. 2 or 4 poles, i.e. 1 or 2 pole-pairs (p_2), output power, P_2 (W).
With these motors connected in cascade, the approximate no-load speeds which can be obtained are given in Table 4.1.

Table 4.1 No-load Speeds for Different Connections

S. No.	*Connection*	*n(rps)*	*Speed* *rps*	*rpm*	*Output*
(a)	**Cumulative cascade**	$\frac{f}{(p_1+p_2)}$			
1.	12 poles, and 2 poles	$\frac{50}{(6+1)}$	7.143	429	$P_1 + P_2$
2.	12 poles, and 4 poles	$\frac{50}{(6+2)}$	6.25	375	$P_1 + P_2$
3.	24 poles, and 2 poles	$\frac{50}{(12+1)}$	3.846	231	$P_1 + P_2$
4.	24 poles, and 4 poles	$\frac{50}{(12+2)}$	3.57	214	$P_1 + P_2$
(b)	**Differential cascade**	$\frac{f}{(p_1-p_2)}$			
5.	12 poles, and 2 poles	$\frac{50}{(6-1)}$	10.0	600	$P_1 - P_2$
6.	12 poles, and 4 poles	$\frac{50}{(6-2)}$	12.5	750	$P_1 - P_2$
7.	24 poles, and 2 poles	$\frac{50}{(12-1)}$	4.545	273	$P_1 - P_2$
8.	24 poles, and 4 poles	$\frac{50}{(12-2)}$	5.0	300	$P_1 - P_2$
(c)	**Main motor only**	$\frac{f}{p_1}$			
9.	12 poles	$\frac{50}{6}$	8.333	500	P_1
10.	24 poles	$\frac{50}{12}$	4.167	250	P_1
(d)	**Auxiliary motor only**	$\frac{f}{p_2}$			
11.	2 poles	$\frac{50}{1}$	50.0	3000	P_2
12.	4 poles	$\frac{50}{2}$	25.0	1500	P_2

PROBLEMS

4.1 A four-pole 50 Hz slip-ring induction motor runs at 1425 rpm at full-load. The rotor resistance per phase is 0.3 Ω. Calculate the additional resistance per phase, which must be inserted in the rotor circuit to lower the speed to 1000 rpm, the torque remaining same as before.

4.2 A three-phase, 415 V, 50 Hz, 74.6 kW, six-pole slip-ring induction motor drives a fan load. The resistance of each rotor phase is equal to 0.2 Ω. The torque required for the fan is proportional to square of the speed. At full-load (74.6 kW) the motor runs at a speed of 970 rpm with slip rings short circuited. The torque-speed relation of the motor may be assumed to be linear from no-load to full-load. Determine the additional resistance required to be inserted into each rotor phase to reduce the speed of the fan to 600 rpm.

4.3 A three-phase, eight-pole, slip-ring induction motor driven by an adjustable speed dc motor is to be used as a variable frequency source at its slip-ring terminals. The stator terminals of the induction motor are connected to a three-phase, 415 V, 50 Hz supply. The frequency is to be controlled between 15 to 100 Hz. Neglect all losses, exciting current and leakage reactance. High-speed operation should be neglected.

(a) Calculate the required speed range of the motor.

(b) If the rotor standstill emf is 170 V, determine the open circuit rotor voltage at 15 and 100 Hz.

(c) Calculate the fractions of the rotor output power supplied by the ac supply and the dc motor at 15 and 100 Hz.

4.4 A three-phase, 415 V, 50 Hz, eight-pole induction motor has the following equivalent circuit parameters referred to the stator:

$$r_1 = 0.07\ \Omega, \quad x_1 = 0.2\ \Omega, \quad r_2' = 0.05\ \Omega, \quad x_2' = 0.2\ \Omega, \quad x_m = 6.5\ \Omega$$

The motor operates on a variable frequency supply maintaining constant Volt/Hz ratio. Find the speed at which the torque is maximum, the supply frequency being 20 Hz.

4.5 A cascade induction motor set consists of two machines. The main motor has eight poles and auxiliary motor six poles. Find the possible synchronous speeds of the set fed from 50 Hz mains.

4.6 A three-phase, 440 V, 50 Hz, 110 kW, 24 pole, 245 rpm slip-ring induction motor has both its stator and rotor windings connected in star. The ratio of the stator to rotor turns is 1.25. The resistance measured between each pair of slip rings is 0.04 Ω. This motor drives a fan, which requires 110 kW at full-load speed of the motor. The torque required to drive the fan varies as square of the speed. What resistance should be connected in series with each slip ring so that the fan will run at 175 rpm? Neglect stator resistance and leakage reactance, and rotational losses.

4.7 A three-phase, star-connected, 415 V, 50 Hz, wound rotor induction motor has the following constants referred to the stator:

$$r_2' = 0.1\ \Omega, \quad x_2' = 0.2\ \Omega$$

The stator resistance, leakage reactance, and magnetizing current may be neglected. The motor speed driving a load, whose torque is constant, is 1440 rpm. Calculate the magnitude and phase of the voltage to be impressed at the slip rings in order that the motor may operate at unity power factor at 1200 rpm.

4.8 A four-pole induction motor and a six-pole induction motor are connected in cumulative cascade, and fed from 50 Hz mains. The frequency in the secondary circuit of the six-pole motor is observed to be 1 Hz. Determine (a) the slip in each machine, and (b) the combined speed of the set.

4.9 A three-phase, 400 V, 50 Hz, four-pole, star-connected wound rotor induction motor has the following parameters referred to the stator:

$$r_1 = 1.5\ \Omega, \qquad x_1 = 3\ \Omega, \qquad r_2' = 2\ \Omega, \qquad x_2' = 3\ \Omega$$

If the motor is used for regenerative braking, determine

(a) the maximum overhauling torque it can hold and the range of speed it can safely operate,

(b) the speed at which it will hold with a load torque of 50 N · m, and

(c) the maximum overhauling torque that the motor can hold as a ratio of maximum overhauling torque without capacitor, if a capacitive reactance of 2 Ω is inserted in each phase of the stator.

CHAPTER 5

HEATING AND POWER RATING OF DRIVE MOTORS

5.1 LOAD DIAGRAM, OVERLOAD CAPACITY

An electric drive operates reliably, if its motor has been selected properly. The right motor for a given application can be found from the load diagram, with possible instantaneous overloads and permissible temperature rise being taken into account. The load diagram shows graphically the variation of torque acting on the electric drive. The motor of the electric drive has to overcome the load torque expressed as a function of time. There are two major types of load diagrams, one for the static or steady state process, and the other for the dynamic process, when the dynamic components of torque are induced by the inertia of the motor and load.

The load diagram of a crane is shown in Fig. 5.1. At time t_1, the hoisting of the load starts and since the load is constant, torque T_L remains constant till time t_2 when the motor is stopped and the pulley is blocked by the clutch. During the time from t_2 to t_3, the load is hanging. But there is no load on the motor as it is opposed by the clutch. From the instant t_3, the lowering process

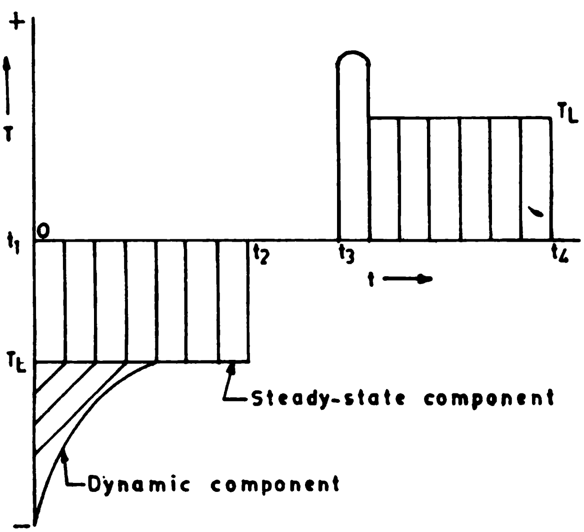


Fig. 5.1 Load diagram of a crane.

starts. The load torque and the motor torque have now the same direction. As the rotation of the shaft reverses, the sign of the load torque also changes.

The motor rating from the standpoint of overloading is considered to be selected properly, if its rated (full-load) torque T_r is governed by

$$T_r > \frac{T_{max}}{\lambda} \tag{5.1}$$

where

T_{max} is the maximum torque required by the driven equipment as evidenced by the load diagram, and

λ is the instantaneous torque overload capacity of the motor.

In dc motors, the maximum value of λ is restricted by the prerequisite of safe commutation, while in ac motors, it is determined by the maximum electromagnetic torque available. The values of λ for different types of motors are given in Table 5.1.

Table 5.1 Values of λ for Different Types of Motors

Type of motor	λ
DC series and compound wound motors (crane)	3.5–4.0
General purpose dc motors	2.5
Squirrel cage and slip-ring induction motors (crane)	2.3–3.4
General purpose squirrel cage and slip-ring induction motors	1.7–2.7
Synchronous motors	2.0–2.7

5.2 INSULATING MATERIALS

The motor rating is considered selected properly from the point of view of heating, if the temperature of insulation under all possible service conditions does not exceed the maximum permissible temperature for the class of insulation used.

The insulating materials used in electrical machines can be grouped into seven classes based on their thermal stability as presented in Table 5.2. The maximum temperatures as listed are evolved for an ambient temperature of 35°C. Therefore, the nameplate rating of a motor is valid for this standard ambient temperature. If the ambient temperature is less than 35°C, the motor can carry a larger load than that stated in its nameplate; if, however, the ambient temperature is greater than 35°C, the load capacity of the motor would be less than its rated capacity.

Table 5.2 Classification of Insulating Materials

Class	*Maximum temperature*	*Insulating material*
Y	90°C	Materials or combination of materials such as cotton, silk or paper without impregnation
A	105°C	The same materials as for Y, but when suitably impregnated or coated or when immersed in a dielectric liquid such as oil

(*Contd.*)

Class	Maximum temperature	Insulating material
E	120°C	Some synthetic organic films
B	130°C	Mica, asbestos, glass, etc. in built up form with organic bonding substances
F	155°C	Mica, asbestos and fiber glass in built up form with synthetic bonding substances and impregnating compounds
H	180°C	Mica, asbestos and fiber glass in built up form with suitable bonding substances such as silicone resin and impregnating compounds
C	Above 180°C	Mica, porcelain, glass and quartz with or without an inorganic binder

5.3 HEATING AND COOLING OF MOTORS

Losses arise in a motor in the process of converting electrical energy into mechanical energy. They appear in the form of heat. Since, the amount of heat produced increases as the motor develops torque to drive the load, the temperature inside the motor would tend towards infinity, if it were not for the fact that some of this heat is dissipated to the atmosphere. After some time has elapsed, the heat created in the motor per unit time is dissipated into the atmosphere during the same period of time and the temperature of the motor reaches its steady state value.

The following assumptions are made in determining the variation of temperature rise (motor temperature minus the ambient temperature) with time:

1. The atmosphere possesses an infinite thermal capacity and, therefore, its temperature does not change due to heat received from a radiating body (the motor).

2. The internal conductivity is infinite and as a result, all parts are at the same temperature.

3. The body is homogeneous, i.e. the conditions for cooling are identical at all points on the surface of the body.

4. The heat losses, the emissivity and the heat capacity do not depend upon temperature.

Denoting by,

Qdt—the heat in calories produced in the motor during time dt.

$A\tau\, dt$—the amount of heat dissipated into the atmosphere in time dt for a temperature rise of τ and emissivity A (cal per sec per °C).

$Cd\tau$—the amount of heat necessary to raise the temperature of the motor having thermal capacity C (cal per °C) through $d\tau$ (°C).

The following heat balance equation is obtained:

$$Qdt = A\tau dt + Cd\tau \tag{5.2}$$

or

$$dt = \frac{Cd\tau}{Q - A\tau}$$

The value of t is computed from the initial condition that at $t = 0$, the initial temperature rise is $\tau = \tau_0$.

Therefore

$$t = -\frac{C}{A}\log_e\left[\frac{Q - A\tau}{Q - A\tau_0}\right] \tag{5.3}$$

The ratio

$$T_H\,(\text{sec}) = \frac{C}{A} = \frac{\text{cal per °C}}{\text{cal per sec per °C}} \tag{5.4}$$

The T_H is called the *thermal time constant*, which is numerically equal to the time for the motor temperature to reach its steady-state value, if no heat is dissipated into the atmosphere.

To find the temperature rise τ as a function of time, we have from Eq. (5.3)

$$-\frac{A}{C}t = \log_e\left[\frac{Q - A\tau}{Q - A\tau_0}\right]$$

or

$$e^{-\frac{A}{C}t} = \left[\frac{Q - A\tau}{Q - A\tau_0}\right]$$

or

$$\tau = \frac{Q}{A}\left(1 - e^{-\frac{A}{C}t}\right) + \tau_0 e^{-\frac{A}{C}t} \tag{5.5}$$

In the steady state ($t = \infty$), therefore, we have

$$\tau = \tau_{ss} = \frac{Q}{A} \tag{5.6}$$

Hence

$$\tau = \tau_{ss}\left(1 - e^{-\frac{t}{T_H}}\right) + \tau_0 e^{-\frac{t}{T_H}} \tag{5.7}$$

If at the outset of the heating process $\tau_0 = 0$, then the equation for the temperature rise takes the form

$$\tau = \tau_{ss}\left(1 - e^{-\frac{t}{T_H}}\right) \tag{5.8}$$

Therefore, the first term in the temperature rise Eq. (5.7) shows the relationship between the temperature rise and time, if the motor is initially cold (heating curve 1, Fig. 5.2).

If no heat is produced, $\tau_{ss} = 0$. Equation (5.7) then takes the form

$$\tau = \tau_0 e^{-\frac{t}{T_H}} \qquad \text{(curve 1, Fig. 5.3)}$$

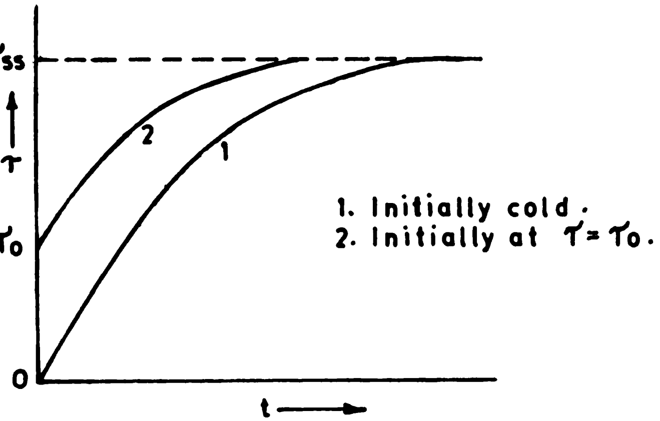


Fig. 5.2 Variation of temperature rise vs. time for heating.

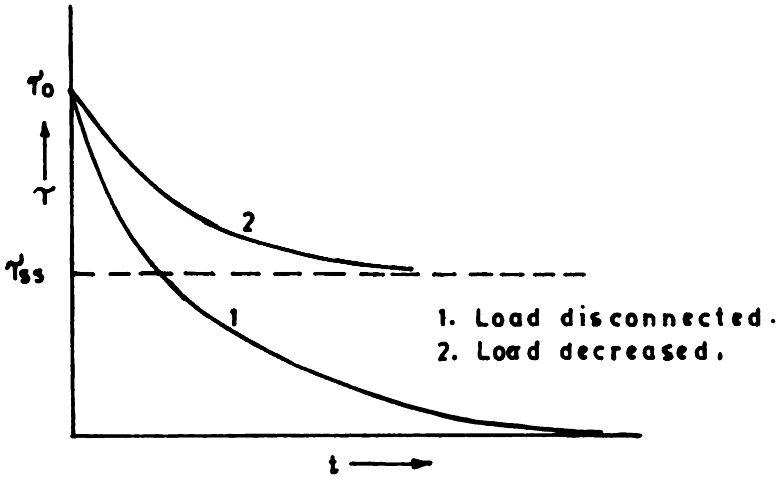


Fig. 5.3 Variation of temperature rise vs. time for cooling.

This is the case after the motor has been disconnected from the circuit. If the load on the motor has been decreased, the steady state temperature rise τ_{ss} in Eq. (5.7) is not equal to zero and the cooling process proceeds according to curve 2 in Fig. 5.3.

From the equations presented above, it can be observed that theoretically the temperature rise τ attains its steady-state value $t_{ss} = Q/A$ only after infinite time has elapsed. However, when $t = 3T_H$, $\tau = 0.95\ \tau_{ss}$ and when $t = 4T_H$, $\tau = 0.98\ \tau_{ss}$. Experiments have shown that practically a motor reaches its steady-state temperature after three to four times T_H, where T_H for squirrel cage self-ventilated motors of open design lies between 11 and 22 minutes and for wound rotor induction motors of open design between 25 to 90 minutes. The emissivity A of enclosed motors is less than that of open motors and, therefore, the time constant $T_H = C/A$ for the former type is greater.

The time constant T_H does not vary with the motor load and as indicated above, it is determined by the parameters C and A, which depend on

(a) G = Weight of the active parts of the machine, kg,

(b) H = Specific heat, cal per kg per °C,

(c) S = Cooling surface, m^2, and

(d) λ = Specific heat dissipation or emissivity, cal per sec per m^2 per °C,

such that,

$$C = G \cdot H, \quad \text{and} \quad A = S \cdot \lambda.$$

The time constant T_H may be evaluated graphically from the tangent to temperature rise vs time curve at $t = 0$ (Fig. 5.4). The segment *ab* is numerically equal to the time constant T_H. Substitution of $t = T_H$ into Eq. (5.8) shows that during this time the temperature rise attains 63.2% of its steady

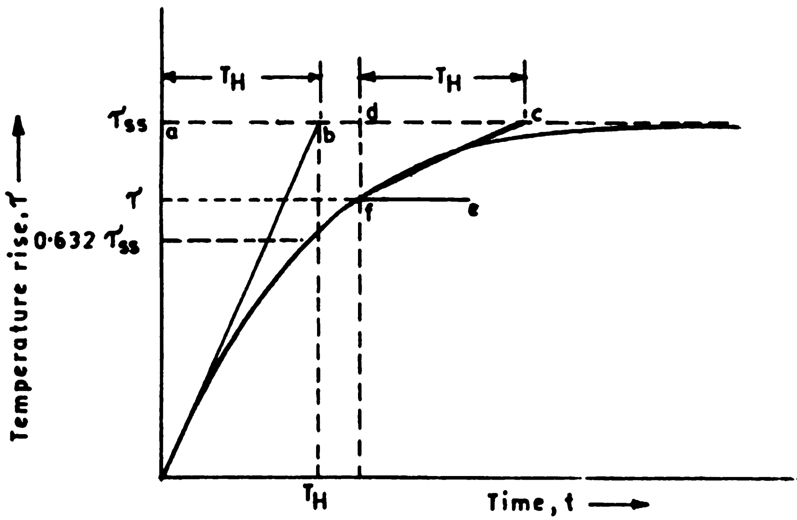


Fig. 5.4 Graphical determination of heating time constant.

state value. It should be noted that the time constant T_H can be obtained graphically by drawing a tangent to the heating curve (Fig. 5.4) not only at the origin, but also at any point *f*, for example. At the point *f*, let the tangent *fc* be drawn. Differentiating Eq. (5.8), we get

$$\frac{d\tau}{dt} = \frac{\tau_{ss}}{T_H} e^{-\frac{t}{T_H}}$$

From Eq. (5.8), we have

$$e^{-\frac{t}{T_H}} = \frac{\tau_{ss} - \tau}{\tau_{ss}}$$

Then

$$\frac{d\tau}{dt} = \frac{\tau_{ss} - \tau}{T_H}$$

or

$$T_H = \frac{\tau_{ss} - \tau}{\frac{d\tau}{dt}} \tag{5.9}$$

Consider now the triangle *fdc*,

$$\tan \angle cfe = \cot \angle fcd = \frac{d\tau}{dt} = \frac{df}{dc} = \frac{\tau_{ss} - \tau}{dc}$$

or

$$dc = \frac{\tau_{ss} - \tau}{\dfrac{d\tau}{dt}}$$

Therefore

$$T_H = dc$$

Determination of steady-state temperature rise

The final (steady-state) temperature rise may be found from a construction of only the initial part of the heating curve, i.e. by measuring the temperature rise increments $\Delta\tau_1$, $\Delta\tau_2$, $\Delta\tau_3$, etc. at time intervals $\Delta\tau_1$, $\Delta\tau_2$, $\Delta\tau_3$, etc. along this part of the curve. The temperature rise increments $\Delta\tau (i = 1, 2, 3, \ldots)$ are plotted to the left as shown in Fig. 5.5. The points, 1, 2, 3, etc. lie on a straight line, which, when extended to its intersection with the ordinate axis, gives the steady-state temperature rise.

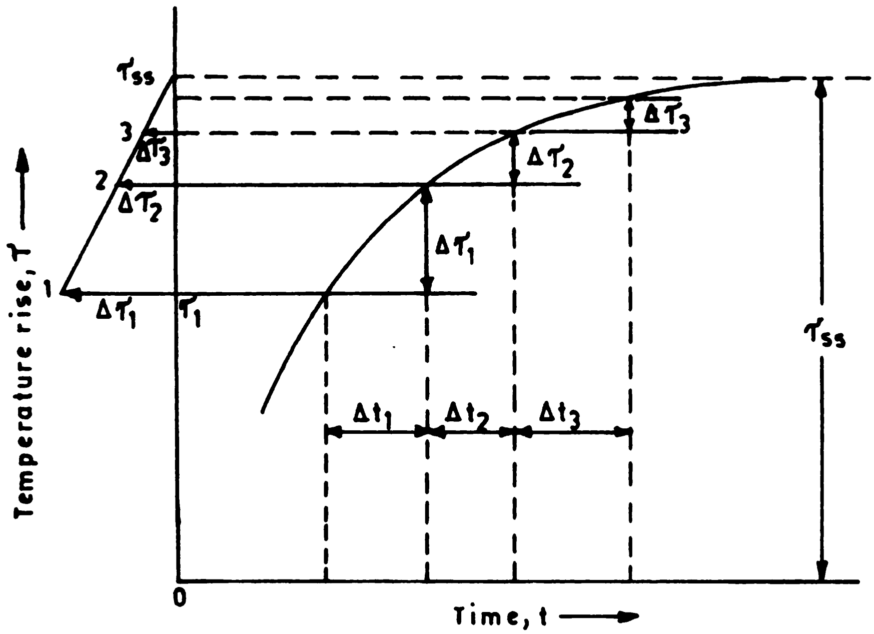


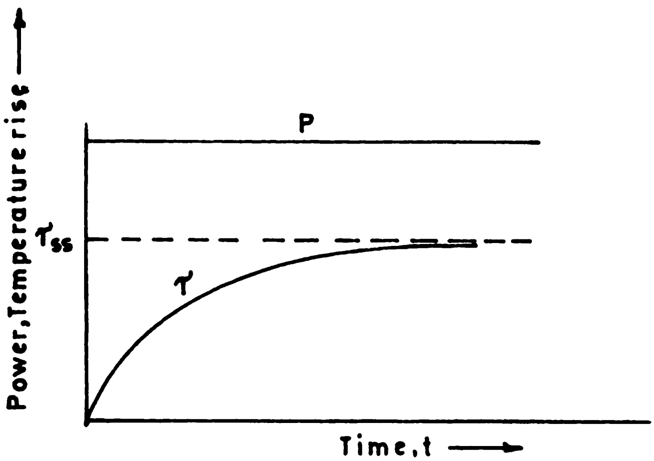
Fig. 5.5 Graphical determination of steady-state temperature rise.

5.4 SERVICE CONDITIONS OF ELECTRIC DRIVES: CONTINUOUS, INTERMITTENT AND SHORT TIME

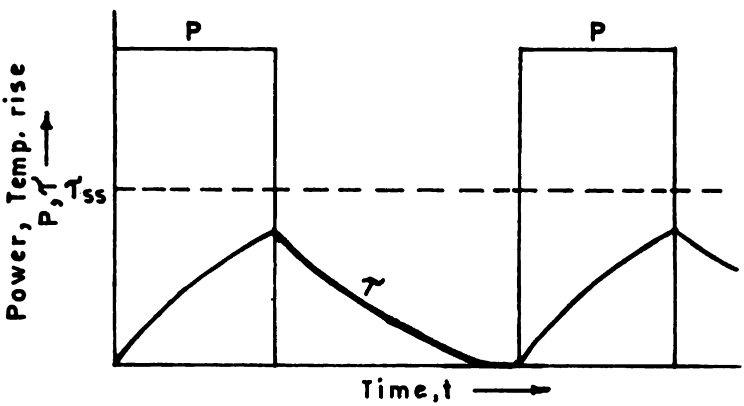
The rating of a motor selected from the standpoint of heating depends on the service conditions under which it is to run. In practice, these conditions are classified, according to duration and nature of the load, into three types, namely continuous, intermittent and short time.

Continuous duty is the term adopted to describe the service condition, wherein the running period is of sufficient duration for the temperature rise to attain its steady state value. It is not necessary that the load be constant throughout. Several types of equipment run continuously with a constant load, e.g. pumps, fans, compressors and conveyors. The load curve for this service condition is shown in Fig. 5.6a. The types of equipment, which run continuously with variable load (Fig. 5.7), are machine tools, for example.

With the machine (motor) on short time duty, the temperature rise does not reach its steady state value during the working period. Moreover, the pause (idle period between consecutive working periods) is of sufficient duration for complete cooling to take place. Auxiliaries for several kinds of machine tools and also for cranes, hoists, etc. remain idle or keep running at no-load for a long time after each working cycle (see Fig. 5.6b).



(a) Continuous duty.



(b) Short-time duty.

Fig. 5.6 Load curves.

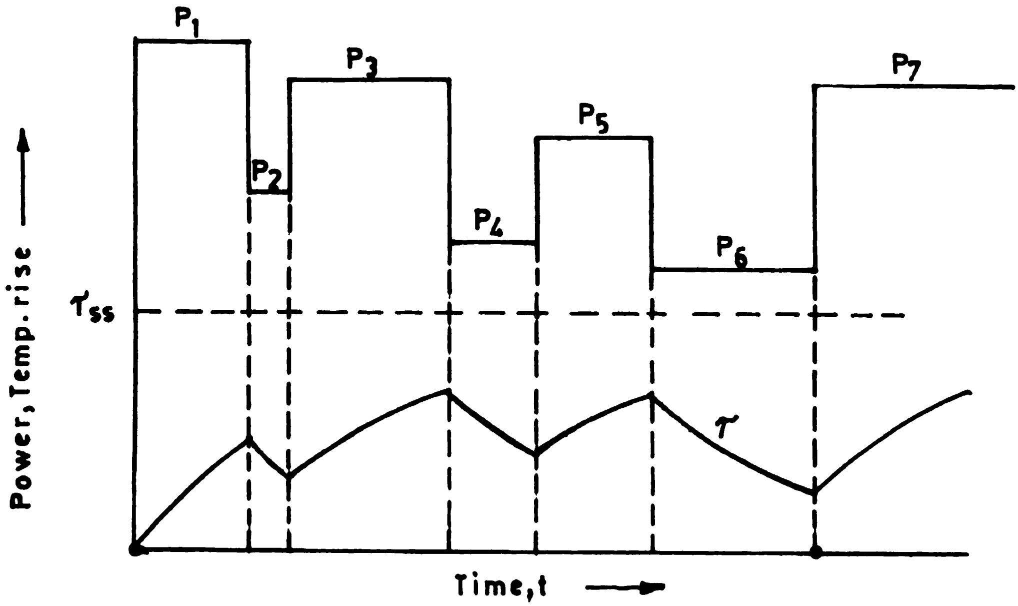


Fig. 5.7 Variable load curve for motors performing continuous duty.

Special types of motors are manufactured for service conditions which require short time ratings. It is the time during which the motors can be run at prescribed loads, stipulated, for example, as 6, 10, 15, 30, 60, 90 and 120 minutes, etc.

Example 5.1

A motor has a thermal heating time constant of 45 minutes. When the motor runs continuously on full-load, its final temperature rise is 80°C.

(a) What would be the temperature rise after 1 hour, if the motor runs continuously on full-load?

(b) If the temperature rise on 1 hr. rating is 80°C, find the maximum steady-state temperature at this rating.

(c) How long will the motor take for its temperature to rise from 50°C to 80°C, if it is working at its 1 hr. rating?

Solution

(a) $T_H = 45$ min., $\tau_{ss} = 80°C$, $t = 1$ hour $= 60$ min.

$$\frac{t}{T_H} = \frac{60}{45} = 1.333$$

$$\tau = \tau_{ss}\left(1 - e^{-\frac{t}{T_H}}\right) = 80(1 - e^{-1.333}) = 58.9°C$$

(b) The heating time constant will remain unchanged. Therefore,

$$80 = \tau_{ss}(1 - e^{-1.333})$$

or

$$\tau_{ss} = \frac{80}{1 - e^{-1.333}} = 108.64°\text{C}$$

(c) $\tau_0 = 50°\text{C}$. Therefore,

$$\tau = \tau_{ss}\left(1 - e^{-\frac{t}{T_H}}\right) + \tau_0 e^{-\frac{t}{T_H}}$$

or

$$80 = 108.64\left(1 - e^{-\frac{t}{45}}\right) + 50e^{-\frac{t}{45}} = 108.64 - 58.64e^{-\frac{t}{45}}$$

or

$$t = 45 \times \log_e\left(\frac{58.64}{108.64 - 80}\right) = 32.25 \text{ min.}$$

For intermittent duty, the running periods and pauses are not long enough to enable the temperature to rise to steady state during the former period and the motor to cool down during the latter period (Fig. 5.8). Several kinds of hoists, machine tools and forging equipments are subjected to intermittent duty. Special motors are manufactured for these service conditions. They are given an equivalent continuous load nameplate rating for a certain duty factor, which is defined as the ratio of the total operating time t_{op} to the total duration of the duty cycle.

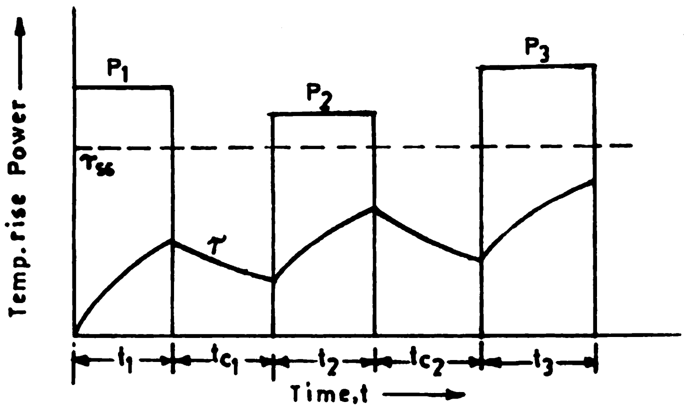


Fig. 5.8 Load curve for motors performing intermittent duty.

That is, with reference to Fig. 5.8, we can write
Duty factor,

$$\varepsilon = \frac{t_{op}}{t} = \frac{(t_1 + t_2)}{[(t_1 + t_2) + (t_{c1} + t_{c2})]} \tag{5.10}$$

where

$$t_c = (t_{c1} + t_{c2})$$

is the total cooling time in the cycle.

Motors with intermittent duty ratings are manufactured for $\varepsilon = 0.15$, 0.25 and 0.40 with their duty cycle, $t = (t_{op} + t_c)$ to last no longer than 10 minutes. Intermittent duty crane motors are designed basically for a duty factor of 0.25; however, they can be used for other values, if the rating is recalculated accordingly.

5.5 SELECTION OF MOTOR POWER CAPACITY

5.5.1 Continuous Duty Constant Load Motor Applications

The nameplate rating P_{nom} of a motor for these conditions should correspond to the shaft power P_o of the driven equipment. After including the losses in the equipment and transmission, it equals

$$P_{nom} = \frac{P_o}{\eta_m \eta_t} \quad (5.11)$$

where η_m and η_t are the efficiencies of the motor and transmission, respectively.

For example, a motor used to drive a pump should have a rating,

$$P_{nom} = \frac{YHQ}{102\eta_p \eta_t} \text{ kW} \quad (5.12)$$

where

Y is the specific gravity of the liquid pumped in kg per m^3,

H is the sum of the static and friction head in m,

Q is the delivery of the pump in m^3 per sec, and

η_p and η_t are the efficiencies of the pump and transmission from motor to pump, respectively.

Similarly, the rating of a fan motor is determined as

$$P_{nom} = \frac{Qh}{102\eta_f \eta_t} \text{ kW} \quad (5.13)$$

where

Q is the delivery of the fan in m^3 per sec,

h is the pressure in kg per m^2, and

η_f and η_t are the efficiencies of the fan and transmission, respectively.

5.5.2 Continuous Duty Variable Load Motor Applications

The rating of a motor for these conditions can be determined either by the method of average losses or by the method of determining equivalent current, torque or power.

Method of average losses

Each motor has its own rated losses Q_{nom}, which at rated load result in an allowable temperature rise. If a motor, selected on the basis of load curve, has average losses Q_{av} equal or close to Q_{nom}, then it is suited to do the job. However, if $Q_{av} > Q_{nom}$, the motor in question will overheat; hence, a motor with a higher rating should be used for the given application. On the other hand, if $Q_{av} < Q_{nom}$, the motor is too large for the purpose for which it has been selected and, therefore, it would only be partially loaded.

The steps involved in the method of average losses are as follows:

(a) An approximate motor rating is chosen from the formula

$$P_{nom} = kP_{av}, \qquad k = 1.1 \text{ to } 1.3 \tag{5.14}$$

$$P_{av} = \frac{P_1 t_1 + P_2 t_2}{t_1 + t_2} \tag{5.15}$$

In the above formula, the duty cycle is taken to have two different loads (P_1 and P_2) instead of six as shown in Fig. 5.7.

(b) The motor losses are calculated separately for each load by the formula,

$$\Delta Q_1 = \frac{P_1}{\eta_1} - P_1$$

$$\Delta Q_2 = \frac{P_2}{\eta_2} - P_2$$

where η_1 and η_2 are the efficiencies at the loads P_1 and P_2, respectively.

The efficiency of the motor at any load x can be calculated from the nominal η_{nom} by using the relationship

$$\eta_x = \frac{1}{1 + \left(\dfrac{1}{\eta_{nom}} - 1\right)\left(\dfrac{x + \dfrac{a}{x}}{1 + a}\right)} \tag{5.16}$$

where

$$a = \left(\frac{q_{const}}{q_{var}}\right)_{nom}$$

= ratio of the motor constant loss to the motor variable loss at rated load

$$Q_{nom} = q_{const} + q_{var}$$

$x = \dfrac{I_x}{I_{nom}}$ = ratio of the motor current for x times full-load to the full-load current

(c) The average losses during the duty cycle are found from the formula,

$$Q_{av} = \frac{\Delta Q_1 t_1 + \Delta Q_2 t_2}{t_1 + t_2} \tag{5.17}$$

If the average losses found from Eq. (5.17) turn out to be less than the rated losses Q_{nom}, the motor can be used for the given application.

This method of average losses gives very accurate results, but it involves many computations.

Equivalent current method

The total losses in a motor consist of constant loss and variable loss depending upon the current, i.e. load. The variable loss can be expressed as $I_{eq}^2 R$, where I_{eq} is the equivalent current producing the same losses in the motor as the actual current in the duty cycle.

$$Q_{av} = q_{const} + q_{var} = q_{const} + I_{eq}^2 R$$

$$= \frac{(q_{const} + I_1^2 R)\cdot t_1 + (q_{const} + I_2^2 R)\cdot t_2 + \ldots + (q_{const} + I_n^2 R)\cdot t_n}{t_1 + t_2 + \ldots + t_n}$$

or

$$I_{eq} = \sqrt{\frac{I_1^2 t_1 + I_2^2 t_2 + \ldots + I_n^2 t_n}{t_1 + t_2 + \ldots + t_n}} \tag{5.18}$$

The rated current of the selected motor should be greater than this equivalent current. It is also necessary to check the current overload capacity of the motor using the formula $\lambda = I_{max}/I_{nom}$, where I_{max} is the maximum current in the load diagram. The proper rating P_{nom} of the motor is found from the catalogue for the rated current and overload capacity obtained above.

Equivalent torque method

The motor rating can also be determined from the load diagram expressed as torque. However, the torque and current of the motor have to be proportional. This proportionality exists in dc motors (separately excited and shunt), where flux is constant and also to a certain extent in induction motors running on their inherent characteristics and on the linear section of the rheostatic characteristics. In this case, the load current is substituted by the load torque in Eq. (5.18), to obtain the expression for the equivalent torque,

$$T_{eq} = \sqrt{\frac{T_1^2 t_1 + T_2^2 t_2}{t_1 + t_2}} \tag{5.19}$$

The rated torque of the motor selected from the catalogue should exceed the equivalent torque and also

$$P_{nom} \geq \frac{T_{eq} N_r}{975} \text{ kW} \tag{5.20}$$

where N_r is the speed in rpm.

It should be noted that the method of equivalent torque is not applied, if the current and torque are not proportional to each other as in dc series motors.

Equivalent power method

This method of determining the rating is applicable to motors running at constant or slightly variable speed, e.g. synchronous, induction and dc shunt and separately excited motors. Since, $P = \omega T$ under the condition, ω = constant, we have, $P = KT$. After substituting P for T into Eq. (5.19), we get

$$P_{eq} = \sqrt{\frac{P_1^2 t_1 + P_2^2 t_2}{t_1 + t_2}} \tag{5.21}$$

The methods discussed above for selecting the rating of a motor from thermal considerations are based on the assumption that the same amount of heat is dissipated per unit time during starting, idling and running.

If the motor runs at variable speed (Fig. 5.9) as during starting or braking or remains standstill during pauses, the conditions for heat dissipation in motors having shaft-mounted fans become poorer. Therefore, the factors for starting (β_{st}), braking (β_{br}) and cooling (β_c) are introduced into Eq. (5.21) for equivalent power, whereupon

$$P_{eq} = \sqrt{\frac{P_{st}^2 t_{st} + P_{br}^2 t_{br} + P_{op}^2 t_{op}}{\beta_{st} t_{st} + t_{op} + \beta_{br} t_{br} + \beta_c t_c}} \tag{5.22}$$

where

$$\beta_c = \frac{T_H}{T_C} \tag{5.23}$$

is the ratio of the heating and cooling time constants.

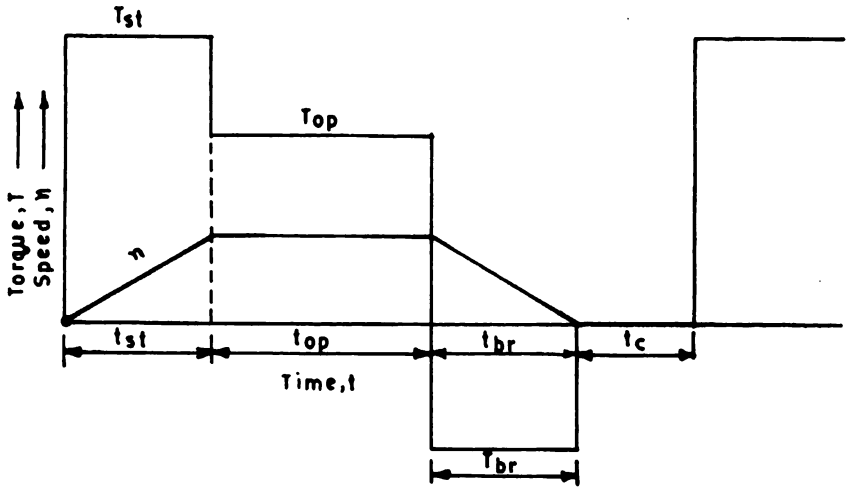


Fig. 5.9 Variations in speed along different sections of the load curve.

The cooling time constant T_C for motors with shaft-mounted fans is 2-3 times larger than T_H, because of the poorer ventilation condition at standstill. In other words, β_c depends upon the design of the motor and varies from 0.25 to 0.35 for self-cooled protected motors, 0.45 to 0.55 for separately cooled enclosed motors. The value of β_c can be increased even to unity by providing separate ventilation.

The factors

$$\beta_{st} = \beta_{br} \frac{(1+\beta_c)}{2} \tag{5.24}$$

If the load curve is irregular, it can be replaced by an equivalent stepped curve (Fig. 5.10), whose steps cover the time intervals. In this case

$$P_{eq} = \sqrt{\frac{\sum_{i=1}^{n} [P_i^2 t]}{nt}} = \sqrt{\frac{\sum_{i=1}^{n} P_i^2}{n}} \tag{5.25}$$

where n is the number of equal time intervals.

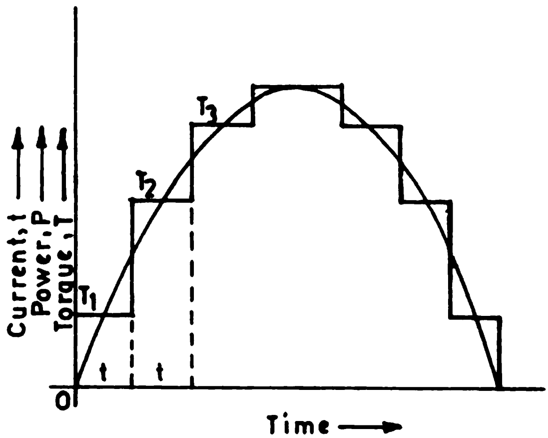


Fig. 5.10 Irregular load curve replaced by an equivalent stepped curve.

The power-time curve is plotted by multiplying the instantaneous speed in rad/s by torque in N · m.

If the power along a certain section of the load curve varies linearly with time from P_1 to P_2 (Fig. 5.11), the equivalent power for this section can be found as shown below.

The ordinate at any instant t is given by

$$h = h_1 + \frac{[(h_2 - h_1)t]}{T}$$

or

$$h^2 = h_1^2 + \frac{2h_1(h_2 - h_1)t}{T} + \frac{(h_2 - h_1)^2 t^2}{T^2}$$

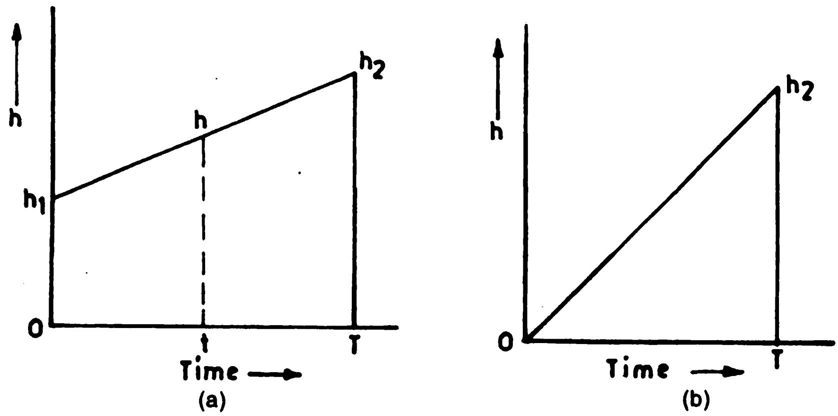


Fig. 5.11 Load varying linearly with time having: (a) An initial value; and (b) Zero initial value.

In terms of power P, where $h_1 = P_1$ and $h_2 = P_2$, we have

$$P_{eq} = \sqrt{\frac{1}{T}\int_0^T P^2 dt} = \sqrt{\frac{1}{T}\int_0^T \left[P_1^2 + \frac{2P_1(P_2 - P_1)t}{T} + \frac{(P_2 - P_1)^2 t^2}{T^2}\right] dt}$$

$$= \sqrt{\frac{P_1^2 + P_1P_2 + P_2^2}{3}} \qquad (5.26a)$$

For the special case in which $P_1 = 0$, we get

$$P_{eq} = \frac{P_2}{\sqrt{3}} \qquad (5.26b)$$

Example 5.2

A motor driving a colliery winding equipment has to deliver a load, having the following characteristics:

(a) Rising uniformly from zero to a maximum of 2000 kW in 20 sec. during the accelerating period,

(b) 1000 kW for 40 sec. during the full-speed period, and

(c) during the deceleration period of 10 sec., when regenerative braking is taking place, the power returned to the supply falls from an initial value of 330 kW to zero.

The interval of decking the next load cycle is 20 sec. (Fig. 5.12). What size of continuously rated motor would be suitable? State the assumptions made.

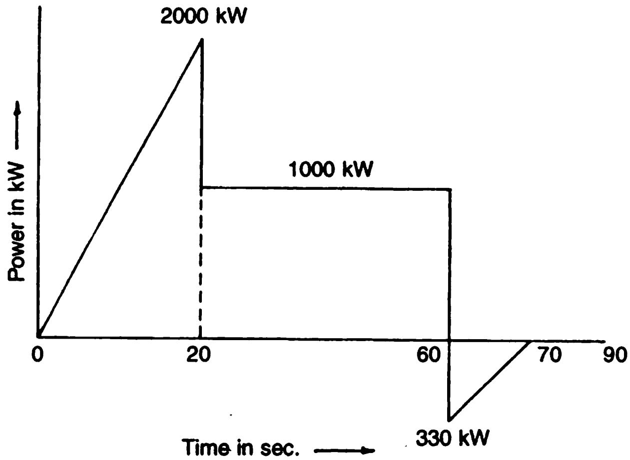


Fig. 5.12 Load (power) diagram of Example 5.2.

Solution

The power load diagram shows only the actual output power. It does not take into account the starting and braking losses as well as other losses. Thus,

$$P_{eq} = \sqrt{\frac{\dfrac{(2000)^2 \times 20}{3} + (1000)^2 \times 40 + \dfrac{(330)^2 \times 10}{3}}{20 + 40 + 10 + 20}} = 863 \text{ kW}$$

Intermittent duty motor applications

The equivalent power for the intermittent load conditions of Fig. 5.8 is

$$P_{eq} = \sqrt{\frac{P_1^2 t_1 + P_2^2 t_2}{(t_1 + t_2) + (t_{c1} + t_{c2})}} \tag{5.27a}$$

Multiplying and dividing the expression under the radical sign by $(t_1 + t_2)$ and substituting by duty factor, we obtain

$$P_{eq} = \sqrt{\left(\frac{P_1^2 t_1 + P_2^2 t_2}{t_1 + t_2}\right) \times \left(\frac{t_1 + t_2}{(t_1 + t_2) + (t_{c1} + t_{c2})}\right)} = P_{eq}(\text{id})\sqrt{\varepsilon} \tag{5.27b}$$

where P_{eq}(id) is the intermittent duty nameplate rating of the motor for a given duty factor ε, defined as

$$\varepsilon = \frac{t_1 + t_2}{(t_1 + t_2) + (t_{c1} + t_{c2})}$$

Since, the duty factor ε evaluated from the load curve may not coincide with standard or catalogue values (ε_{st} = 0.15, 0.25, 0.40, 0.60) the intermittent duty motor rating, P_{eq}(id) obtained from the above formula should be recalculated for one of the standard duty factors. To do this, it is assumed that the equivalent power P_{eq} remains constant and, therefore

$$P_{eq} = P_{eq}(\text{id})\sqrt{\varepsilon} = P_{eq}(\text{st})\sqrt{\varepsilon_{st}} \tag{5.28}$$

or

$$P_{eq}(\text{st}) = P_{eq}(\text{id})\sqrt{\frac{\varepsilon}{\varepsilon_{st}}} \tag{5.29}$$

Example 5.3

Select the motor for driving the equipment which has the load (torque) curve of Fig. 5.13:

(a) For the first 10 seconds the torque is constant and equal to 41 kg · m;

(b) For the next 30 seconds the torque drops linearly with time from 38 kg · m to 17 kg · m;

(c) For the last 46 seconds the torque is constant and equal to 8 kg · m.

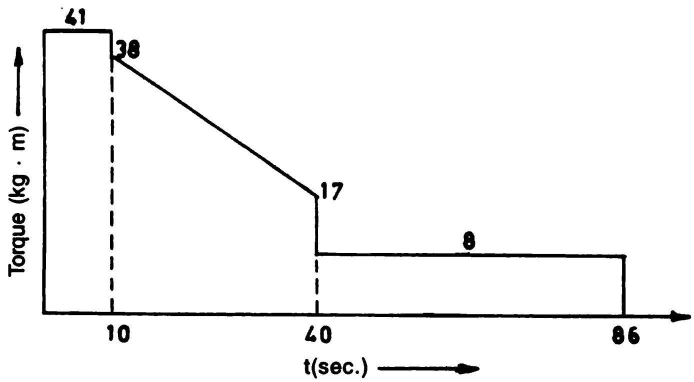


Fig. 5.13 Load (torque) curve of Example 5.3.

Solution

$$P_{eq} = \sqrt{\frac{41^2 \times 10 + \dfrac{(38^2 + (38 \times 17) + 17^2 \times 30}{3} + 8^2 \times 46}{10 + 30 + 46}} = 22.5 \text{ kg} \cdot \text{m}$$

The torque overload capacity is taken equal to

$$\lambda = \frac{T_{max}}{T_{nom}} = 1.7$$

Therefore

$$T_{nom} = \frac{41}{1.7} = 24 \text{ kg} \cdot \text{m}$$

which is greater than 22.5 kg · m.

The motor should have power rating

$$P_{nom} = \frac{T_{nom} N_{nom}}{975} = \frac{24 \times 730}{975} = 18 \text{ kW}$$

A standard 20 kW motor at 730 rpm can be used for this application having maximum to rated torque ratio equal to 1.7 and starting to rated torque ratio equal to 1.1.

Example 5.4

Plot the load curve and select the proper motor for the following intermittent duty:

(a) P_1 = 35 kW for t_1 = 3 sec.
(b) P_2 = 17 kW for t_2 = 20 sec.
(c) P_3 = 35 kW for t_3 = 2 sec.
(d) P_4 = 13 kW for t_4 = 15 sec.

Between the operating periods (b) and (c), there is a pause (P = 0) of t_{c1} = 37 sec.; at the end of the cycle, there is another pause t_{c2} of 40 sec.

Solution

The load curve is shown in Fig. 5.14 below.

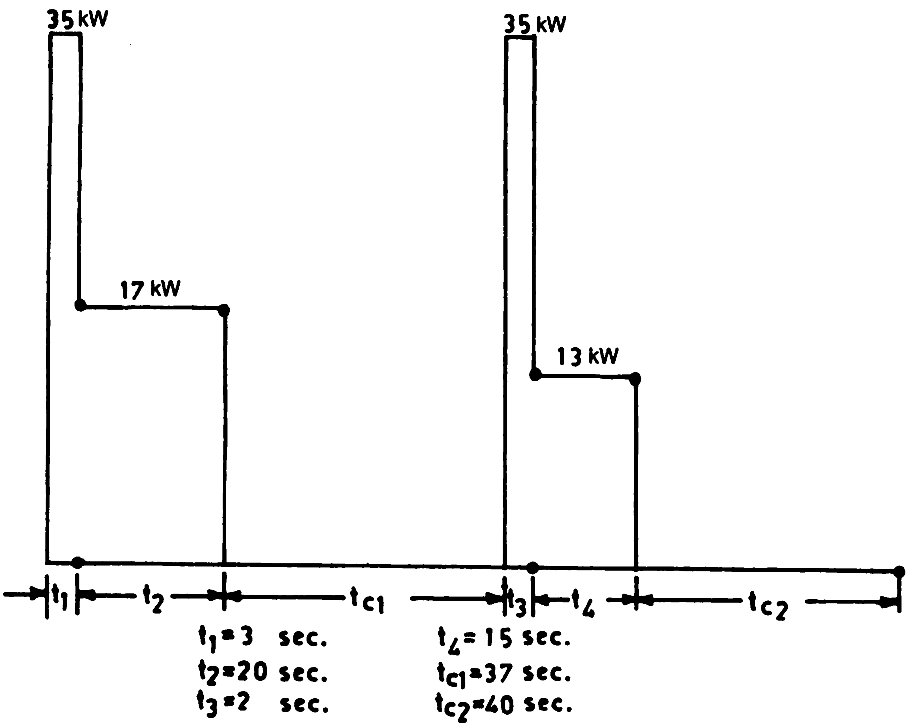


Fig. 5.14 Load curve of Example 5.4.

Duty factor,

$$\varepsilon = \frac{t_{op}}{t_{op} + t_c} = \frac{3 + 20 + 2 + 15}{(3 + 20 + 2 + 15) + (37 + 40)} = 0.34$$

The equivalent continuous load power amounts to

$$P_{eq} = \sqrt{\frac{(35^2 \times 3) + (17^2 \times 20 + 35^2 \times 2) + (13^2 \times 15)}{3 + 20 + 2 + 15}} = 19 \text{ kW}$$

The motor rating for the standard duty factor of 0.40 is

$$P_{eq}(\text{st}) = P_{eq}(\text{id})\sqrt{\frac{\varepsilon}{\varepsilon_{st}}} = 19\sqrt{\frac{0.34}{0.4}} = 17.5 \text{ kW}$$

So, a three-phase, 17.5 kW, 725 rpm, cage induction motor can be used.

Example 5.5

Select a motor to be run at 970 rpm for the following intermittent duty:

(a) $P_1 = 4$ kW for $t_1 = 2$ sec.

(b) $P_2 = 8$ kW for $t_2 = 3$ sec.

(c) $P_3 = 4$ kW for $t_3 = 2$ sec.

(d) $P_4 = 8$ kW for $t_4 = 3$ sec.

(e) $P_5 = 0$ for $t_5 = 5$ sec.

Solution

The equivalent power is equal to

$$P_{eq} = \sqrt{\frac{(4^2 \times 2) + (8^2 \times 3 + 4^2 \times 2) + 8^2 \times 3}{(2 + 3 + 2 + 3) + 5}} = 5.465 \text{ kW}$$

Duty factor,

$$\varepsilon = \frac{2 + 3 + 2 + 3}{(2 + 3 + 2 + 3) + 5} = 0.667$$

$$P_{eq}(\text{id}) = P_{eq}\sqrt{\varepsilon} = 5.465 \times \sqrt{0.667} = 6.7 \text{ kW}$$

Recalculating the rating for a standard duty factor ($\varepsilon = 0.6$), we have

$$P_{eq}(\text{st}) = P_{eq}(\text{id})\sqrt{\frac{\varepsilon}{\varepsilon_{st}}} = 6.7 \times \sqrt{\frac{0.667}{0.6}} = 7.06 \text{ kW}$$

A standard three-phase induction motor of rating 7.5 kW, 970 rpm having maximum to rated torque ratio equal to 1.8 and starting to rated torque ratio equal to 1.1 may be used.

Temperature rise

Let τ_{h1}, τ_{h2}, τ_{h3}, . . . be the respective rises in temperature during the periods when the load is on, and τ_{c1}, τ_{c2}, τ_{c3}, . . . be those when the load is off (Fig. 5.15). T_H and T_C are the heating and cooling

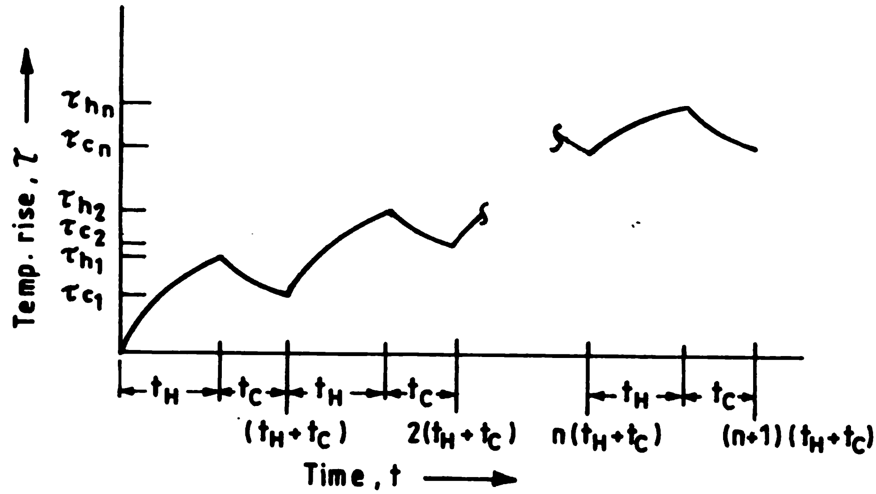


Fig. 5.15 Temperature rises of motor performing intermittent duty.

time constants respectively, and τ'_{ss} is the permissible temperature rise. Let t_h and t_c be the durations of heating and cooling, respectively. Then

$$\tau_{h1} = \tau'_{ss}\left(1 - e^{-\frac{t_h}{T_H}}\right)$$

$$\tau_{c1} = \tau_{h1}e^{-\frac{t_c}{T_C}}$$

Putting, $x = -\dfrac{t_h}{T_H}$ and $y = -\dfrac{t_c}{T_C}$, we get

$$t_{h1} = \tau'_{ss}(1 - e^x)$$

$$\tau_{c1} = \tau_{h1}e^y = \tau'_{ss}(1 - e^x)e^y$$

$$\tau_{h2} = \tau'_{ss}(1 - e^x) + \tau'_{c1}e^x = \tau'_{ss}(1 - e^x) + \tau'_{ss}(1 - e^x)e^xe^y = \tau'_{ss}(1 - e^x)(1 + e^{(x+y)})$$

$$\tau_{c2} = t_{h2}e = \tau'_{ss}(1 - e^x)(1 + e^{(x+y)})$$

$$\tau_{h3} = \tau'_{ss}(1 - e^x) + \tau_{c2}e^x = \tau'_{ss}(1 - e^x) + \tau'_{ss}e^{(x+y)}(1 - e^x)(1 + e^{(x+y)})$$

$$= \tau'_{ss}(1 - e^{x})(1 + e^{(x+y)} + e^{2(x+y)})$$

$$\vdots$$

$$\tau_{hn} = \tau'_{ss}(1 - e^{x})(1 + e^{(x+y)} + e^{2(x+y)} + \ldots + e^{(n-1)(x+y)})$$

$$= \tau'_{ss}\frac{(1-e^{x})(1-e^{n(x+y)})}{1-e^{(x+y)}} \tag{5.30}$$

As $n \to \infty$, $e^{n(x+y)} \to 0$, as both x and y are negative. Then

$$\tau_{hn} = \tau'_{ss}\frac{1-e^{x}}{1-e^{(x+y)}} \tag{5.31}$$

τ_{hn} should not exceed τ'_{ss}, the maximum temperature rise corresponding to the continuous rated load P_{nom}. So

$$\tau_{ss} = \tau'_{ss}\frac{1-e^{x}}{1-e^{(x+y)}} = \tau'_{ss}\frac{1-e^{-\frac{t_h}{T_H}}}{1-e^{-\left(\frac{t_h}{T_H}+\frac{t_c}{T_C}\right)}} \tag{5.32}$$

W_x and W_L are assumed to be the losses corresponding to P_x and P_{nom}, respectively. The losses at rated load,

$$W_L = \text{constant loss} + \text{copper loss} = W_{cont.} + W_{cu} = kW_{cu} + W_{cu} = (k+1)W_{cu} \tag{5.33}$$

where k is a constant.

$$W_x = W_{const.} + W_{cu}\left(\frac{P_x}{P_{nom}}\right)^2 = kW_{cu} + W_{cu}\left(\frac{P_x}{P_{nom}}\right)^2 = W_{cu}\left[k + \left(\frac{P_x}{P_{nom}}\right)^2\right] \tag{5.34}$$

Then

$$\frac{W_x}{W_L} = \frac{\tau'_{ss}}{\tau_{ss}} = \frac{1-e^{-\left(\frac{t_h}{T_H}+\frac{t_c}{T_C}\right)}}{1-e^{-\frac{t_h}{T_H}}} = \frac{k+\left(\frac{P_x}{P_{nom}}\right)^2}{k+1}$$

and

$$\frac{P_x}{P_{nom}} = \sqrt{\frac{(k+1)\left(1-e^{-\left(\frac{t_h}{T_H}+\frac{t_c}{T_C}\right)}\right)}{1-e^{-\frac{t_h}{T_H}}} - k} \tag{5.35}$$

Example 5.6

A 100 kW motor, having rated temperature rise of 80°C, has full-load efficiency of 90% and the maximum efficiency occurs at 80% full-load. It has thermal time constants of 60 minutes and 75 minutes. It is cyclically loaded, 120% of full-load for one hour and 50% of full-load for the next hour. Find the temperature rise after four hours.

Solution

$$k = (0.8)^2 = 0.64, \qquad \eta = 0.9, \qquad P_o(\text{output}) = 100 \text{ kW}$$

$$W_L \text{ (Full-load loss)} = P_o\left(\frac{1}{\eta} - 1\right) = 100\left(\frac{1}{0.9} - 1\right) = 11.11 \text{ kW}$$

$$W_L = (k + 1)W_{cu} = 1.64\ W_{cu}$$

Then

$$W_{cu} = \frac{11.11}{1.64} = 6.775 \text{ kW}$$

$$W_{const.} = W_L - W_{cu} = 11.11 - 6.775 = 4.335 \text{ kW}$$

At 120% of full-load, the permissible (steady-state) temperature rise is τ'_{ss}. At this load

$$W_x = W_{cu}\left[k + \left(\frac{P_x}{P_{nom}}\right)^2\right] = 6.775[0.64 + (1.2)^2] = 14.092 \text{ kW}$$

$$\frac{\tau'_{ss}}{\tau_{ss}} = \frac{W_x}{W_L} = \frac{14.092}{11.11} = 1.268$$

$$\tau'_{ss} = 1.268\tau_{ss} = 1.268 \times 80 = 101.46°\text{C}$$

At 50% of full-load, the steady-state temperature is τ''_{ss}. At this load

$$W_x = 6.775[0.64 + (0.5)^2] = 6.03 \text{ kW}$$

$$\frac{\tau''_{ss}}{\tau_{ss}} = \frac{W_x}{W_L} = \frac{6.03}{11.11} = 0.543$$

$$\tau''_{ss} = 0.543 \times 80 = 43.42°\text{C}$$

$$\frac{t_h}{T_H} = \frac{60}{60} = 1.0$$

$$\tau_{h1} = \text{Temperature rise after 1 hour} = \tau'_{ss}\left(1 - e^{-\frac{t_h}{T_H}}\right) = 101.46(1 - e^{-1}) = 64.135°\text{C}$$

τ_{c1} = Temperature rise after 2 hours

$$= \tau''_{ss}\left(1-e^{-\frac{t_h}{T_H}}\right)+\tau_{h1}e^{-\frac{t_h}{T_H}} = 43.42(1 - e^{-1}) + 64.135e^{-1} = 27.45 + 23.6 = 51.05°\text{C}$$

τ_{h2} = Temperature rise after 3 hours

$$= \tau'_{ss}\left(1-e^{-\frac{t_h}{T_H}}\right)+\tau_{c1}e^{-\frac{t_h}{T_H}} = 101.46(1 - e^{-1}) + 51.05e^{-1} = 64.135 + 18.78 = 82.195°\text{C}$$

τ_{c2} = Temperature rise after 4 hours

$$= \tau''_{ss}\left(1-e^{-\frac{t_h}{T_H}}\right)+\tau_{h2}e^{-\frac{t_h}{T_H}} = 43.42(1 - e^{-1}) + 82.915e^{-1} = 27.45 + 30.5 = 57.95°\text{C}$$

Note: The case of cooling is not taken into account here, as load is only decreased in the second-half (1 hr) of the two-hour period. In the subsection '**Temperature rise**', cooling is considered in the latter period, when the machine is on no-load.

Short time duty motor applications

The rating of a motor for short time duty should be so selected such that its temperature rise during the operating period t_{op} attains the permissible temperature rise for continuous service conditions. The thermal capacity of the motor is then completely utilized.

Let us consider the heating conditions for continuous and short time duties. The permissible temperature rise for continuous duty is

$$\tau = \frac{Q}{A} \tag{5.36}$$

For short time duty,

$$\tau_k = \frac{Q_k}{A}\left(1-e^{-\frac{t_k}{T_H}}\right) \tag{5.37}$$

Q and Q_k are the heat losses in the motor under the continuous, and short time duty conditions, respectively.

After equating Eqs. (5.36) and (5.37), we obtain

$$\frac{Q}{A} = \frac{Q_k}{A}\left(1-e^{-\frac{t_k}{T_H}}\right)$$

or

$$\frac{Q_k}{Q} = \frac{1}{1 - e^{-\frac{t_k}{T_H}}} = K_T \tag{5.38}$$

where K_T is called the *thermal overload factor* being equal to the ratio of the losses under short time and continuous service conditions. In order to find out how the duration t_k for short time operation varies with thermal overload factor K_T, Eq. (5.38) is rewritten as follows:

$$Q_k - Q = Q_k e^{-\frac{t_k}{T_H}}$$

or

$$\log_e\left(\frac{Q_k}{Q_k - Q}\right) = \frac{t_k}{T_H}$$

or

$$t_k = T_H \log_e\left(\frac{K_T}{K_T - 1}\right) \tag{5.39}$$

Since heat losses are proportional to the square of current or power, we have

$$K_T = \frac{Q_k}{Q} = \left(\frac{P_k}{P_{\text{nom}}}\right)^2 = \frac{1}{1 - e^{-\frac{t_k}{T_H}}}$$

where P_k is the permissible short time rating of the motor from the standpoint of heating and P_{nom} is the continuous duty motor rating. Therefore

$$P_k = P_{\text{nom}}\sqrt{K_T} \tag{5.40}$$

Besides the thermal overload factor, the mechanical overload factor for short time duty,

$$K_m = \frac{P_k}{P_{\text{nom}}}$$

is also important, since it characterizes the overload capacity of the motor. The two overload factors evidently are related as follows:

$$K_m = \sqrt{K_T} \tag{5.41}$$

The curve showing how K_m varies with operating time t_{op} expressed in pu form,

$$t_{op}(\text{pu}) = \frac{t_{op}}{T_H}$$

is presented in Fig. 5.16. It is observed that for $t_{op}/T_H < 0.35$, the mechanical overload factor is $K_m > 2.5$. In this case, the motor rating should be selected for overload capacity rather than for heating conditions. However, this leads to oversizing (that is, too large a motor) with low efficiency and power factor.

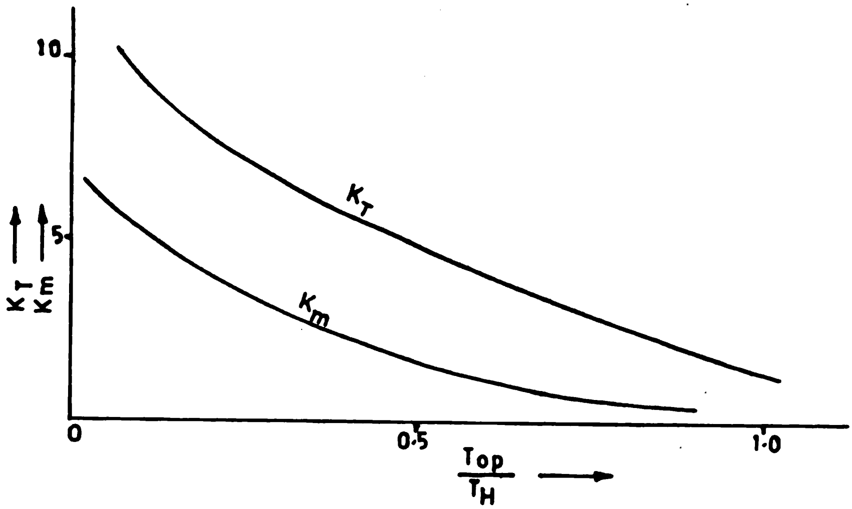


Fig. 5.16 Variation of overload factors with operating time.

Since it is not advisable to use continuous duty motors for such short time service, special motors with specified operating periods are manufactured, e.g. half-hour rating, one-hour rating, etc. If a continuous duty motor is used for short time service with load having two torque levels (T_1 and T_2), the equivalent torque equals

$$T_{eq} = \sqrt{\frac{T_1^2 t_1 + T_2^2 t_2}{t_1 + t_2}} \tag{5.42}$$

The motor rating from heating viewpoint is selected in accordance with T_{eq}. In doing so, the catalogue value of torque overload capacity, $T_{max}/T_{nom} = P_{max}/P_{nom}$ should be greater than P'_{max}/P_{nom}, where P'_{max} is the maximum ordinate of the curve.

The following trial and error method can also be used for selecting short time duty motors. A certain rating of the motor is selected beforehand and heating and cooling time constants are determined. A pattern is made of heating and cooling processes, which is laid on the load curve with heating curve constructed during running period and cooling curve during idling period. If the final temperature does not exceed its permissible value, the motor has been properly selected from its heating standpoint.

Example 5.7

A Q kW electric motor with improved synthetic insulation has a rated temperature rise of 80°C and its efficiency is maximum at 85% of its rated load. If the heating and cooling time constants are

2 hours and 3 hours respectively, what maximum load can the motors carry for 15 minutes followed by a rest of 75 minutes?

Solution

τ_{ss} = Final temperature for continuous rated load

τ'_{ss} = Maximum permissible temperature rise for short time duty

$$\tau_{ss} = \tau'_{ss}\left(1 - e^{-\frac{t}{T_H}}\right)$$

where t = time to attain τ_{ss}.

$$\frac{\tau'_{ss}}{\tau_{ss}} = \frac{W_x}{W_L} = \frac{1}{1 - e^{-\frac{t}{T_H}}}$$

Now

Q_x = Quarter-hour (15 min.) rating, $K = (0.85)^2 = 0.7225$

t_h = 15 min., T_H = 120 min., t_c = 75 min., T_C = 180 min.

$$\frac{t_h}{T_H} = \frac{15}{120} = 0.125, \quad \frac{t_c}{T_C} = \frac{75}{180} = 0.4167$$

$$\frac{Q_x}{Q} = \sqrt{\frac{1 + 0.7225}{1 - e^{-0.125}} - 0.7225} = 3.733$$

$$\frac{\tau'_{ss}}{\tau_{ss}} = \frac{W_x}{W_L} = \frac{K + \left(\frac{Q_x}{Q}\right)^2}{K + 1} = 8.51$$

Temperature rise after 15 minutes

$$\tau_{15} = \tau'_{ss}\left(1 - e^{-\frac{t_h}{T_H}}\right) = (8.51 \times 80)(1 - e^{-0.125}) = 80°\text{C} = \tau_{ss}$$

Temperature after a rest of 75 minutes

$$\tau_{75} = \tau_{ss} e^{-\frac{t_c}{T_C}} = 80 e^{-0.4167} = 52.74°\text{C}$$

It means that the motor does not cool down in 75 minutes. Therefore, the rating has to be reduced.

First trial. It is assumed that $\frac{Q_x}{Q} = 1.0$. $\frac{W_x}{W_L} = 1.0$, corresponding to $\tau'_{ss} = 80°C$.

Temperature rise after 15 minutes = $80(1 - e^{-0.125}) = 9.4°C$

Temperature after a rest of 75 minutes = $9.4e^{-0.4167} = 6.2°C$

Second trial. It is assumed that $\frac{Q_x}{Q} = 0.5$, corresponding to

$$\frac{W_x}{W_L} = \frac{K + \left(\frac{Q_x}{Q}\right)^2}{K + 1} = 0.5646 \text{ and } \tau'_{ss} = 80 \times 0.5646 = 45.167°C$$

Temperature rise after 15 minutes = $45.167(1 - e^{-0.125}) = 5.31°C$

Temperature after a rest of 75 minutes = $5.31e^{-0.4167} = 3.5°C$

It shows that $\frac{Q_x}{Q}$ may be taken equal to 1.0.

5.6 OPERATION OF ELECTRIC DRIVES INCORPORATING FLYWHEEL UNDER SHOCK LOADING CONDITIONS

Some machines like rolling mill, forging machine, electric press, etc. undergo shock loading during their operation. A simplified load diagram is shown in Fig. 5.17, as an example.

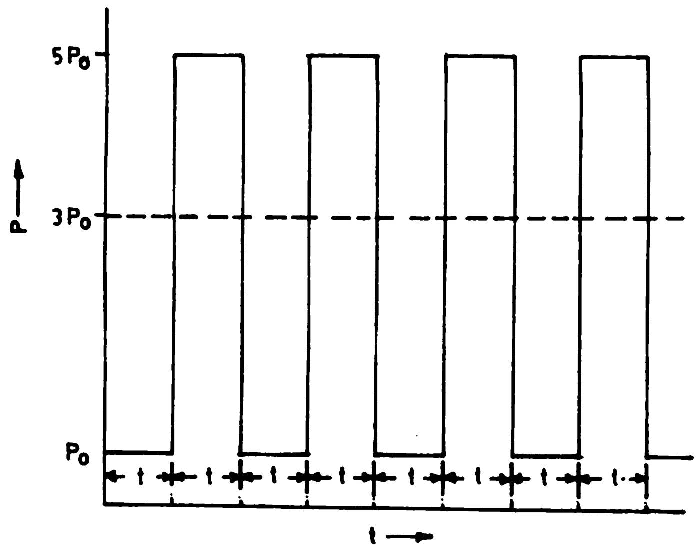


Fig. 5.17 Simplified load diagram of a machine under shock loading.

It has been assumed that sudden loading alternates with an idling period of equal duration. In such drives the losses are high and hence, efficiency is low. The size of the motor has to be increased to bear overloading. The problem can be overcome by fitting a flywheel to the shaft of the motor as the load is then shared by the flywheel. When a load is applied, the speed drops from ω_1 to ω_2 and the stored energy released by the flywheel to share the load with the motor is equal to $1/2J[(\omega_1)^2 - (\omega_2)^2]$, being the flywheel inertia. When the load is off following the peak, the speed rises and the stored energy in the flywheel increases to a new value depending upon the new speed. In this way the load on the motor is smoothed out, thus reducing the losses. This is called the *load equalization*. This can be illustrated with reference to the load diagram of Fig. 5.17. The variable losses are proportional to $(\text{current})^2$, i.e. $(\text{power})^2$. For one cycle, the variable losses are

$$c5P_0^2t + cP_0^2t = 26cP_0^2t$$

When the load on the motor is smoothed out, the average losses are

$$c3P_0^2(2t) = 18cP_0^2t$$

The energy saved per cycle is about 30%.

When the flywheel is incorporated, the motor selected can be of lower rating with lower overload capacity. Now, let us consider the load diagram of a rolling mill to illustrate how the load is shared between the motor and the flywheel, thus lowering the rating and the overload capacity of the motor.

The rolling mill operation consists of a number of passes per schedule. The rolling torque, i.e. load torque remains constant during a pass, but it varies from pass to pass. To find the division of load between the motor and the flywheel, a portion of the load diagram may be considered (Fig. 5.18).

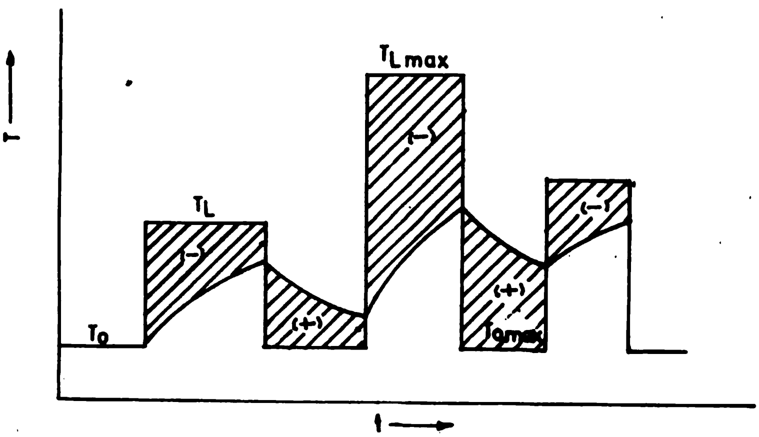


Fig. 5.18 Load sharing between the motor and the flywheel of a rolling mill.

With reference to Fig. 5.18, the torque equation of the motor during a pass, when the metal passes between the rolls of the mill at constant load torque T_L, can be written as

$$T = T_L\left(1 - e^{-\frac{t}{T_{em}}}\right) + T_0 e^{-\frac{t}{T_{em}}} \tag{5.43}$$

When the metal is out of the rolls, we have

$$T = T_0\left(1 - e^{-\frac{t}{T_{em}}}\right) + T_L e^{-\frac{t}{T_{em}}} \tag{5.44}$$

where T_{em} is the electromechanical time constant.

In Fig. 5.18, the shaded area bearing the (–) sign gives the energy supplied by the flywheel to the shaft and the area bearing the (+) sign gives the energy fed to the flywheel.

The size of the flywheel to be fitted to the shaft of the motor is determined based on the fact that the motor has to supply maximum torque equal to λT_{nom}. The maximum value of load torque in the load diagram may be considered for calculation of the size of the flywheel.

$$T = \lambda T_{nom} = (T_L)_{max}\left(1 - e^{-\frac{t}{T_{em}}}\right) + (T_0)_{max} e^{-\frac{t}{T_{em}}} \tag{5.45}$$

where

$(T_0)_{max}$ = Motor torque at the beginning of the maximum torque load period.

Solving Eq. (5.45), we get

$$\frac{t_k}{T_{em}} = \log_e\left[\frac{(T_L)_{max} - (T_0)_{max}}{(T_L)_{max} - \lambda T_{nom}}\right] \tag{5.46}$$

Now

$$T_{em} = \frac{J\omega_0\, s_{nom}}{T_{nom}}$$

The expression of the required moment of inertia of the flywheel is

$$J = \frac{T_{nom} t_k}{\omega_0\, s_{nom} \log_e\left[\frac{(T_L)_{max} - (T_0)_{max}}{(T_L)_{max} - \lambda T_{nom}}\right]} \tag{5.47}$$

It is seen from the above equation that the inertia of the flywheel is decreased, by increasing s_{nom}. It may be borne in mind that a higher value of s_{nom} means higher losses and lower speed. Thus, s_{nom} must lie between 10–15%.

PROBLEMS

5.1 A motor driving mining equipment delivers a load having the following load cycle:

40 kW for 10 min.

No load for 5 min.

20 kW for 10 min.

No load for 5 min.

Estimate the suitable size of a continuously rated motor for the purpose.

5.2 An electric motor is subjected to a load-torque variation as given below:

240 N · m for 20 min.

140 N · m for 10 min.

300 N · m for 10 min.

200 N · m for 20 min.

If the speed of the motor is 720 rpm, find the power rating of the motor.

5.3 A motor works on a three-minute load cycle of the following schedule:

Load rising uniformly from 0 to 800 kW from 0 to 25 sec.

Load remaining constant at 450 kW from 25 to 145 sec.

Power returned falling from 15 kW to 0 from 145 to 160 sec.

Motor remaining stationary from 160 to 180 sec.

What size of a continuously rated motor would be suitable, assuming that the rating depends on the rms value of the load?

5.4 A motor is subjected to an intermittent periodic duty load of xP_{nom} kW with duty factor of ε, the period of load being t_1 and the load cycle being repeated indefinitely. The heating and cooling time constants may be assumed to be equal and the ratio of constant losses to full-load copper loss may be taken equal to a. Derive the expression for the overload factor x.

5.5 An induction motor has a final steady-state temperature rise of 45°C when running at full-load. Copper loss at full-load is 1.2 times the constant losses. The heating time constant is 80 minutes. Calculate the half-hour rating of the motor for the same temperature rise of 45°C.

5.6 After operating for 30 minutes on full-load, the temperature of a motor is 55°C and after 60 minutes on the same load, the temperature becomes 70°C. The ambient temperature may be assumed to be 30°C. Calculate the (a) final temperature, and (b) heating time constant. How long will it take for temperature to rise to 5/6th of its final steady state value, after the start of the motor?

5.7 A 50 kW, three-phase, 440 V, 50 Hz, 1440 rpm, cage induction motor has constant loss to variable loss at full-load in the proportion of 1:3. Its rated temperature rise is 55°C and its heating and cooling time constants are 40 and 60 minutes, respectively. Find

(a) the half-hour rating, and

(b) the intermittent rating, if periodic loads of half-hour duration are applied at an interval of half-hour.

5.8 A motor has heating time constant of 70 minutes and cooling time constant of 80 minutes. When run continuously on full-load of 200 kW, the final temperature rise is 40°C. The motor is driving an intermittent load, remaining 10 minutes on load followed by a 15 minutes shut down period. What is the maximum value of load it can take during the on-load period without the maximum temperature rise exceeding 40°C? Assume losses to be proportional to the square of the load.

5.9 A motor has heating time constant of 70 minutes and cooling time constant when stationary of 90 minutes. When run continuously on full-load of 20 kW, the final temperature rise is 40°C. Assume losses to be proportional to the square of the load.

(a) What load can the motor deliver for 10 minutes, if followed by a shutdown period long enough for cooling, without temperature rise exceeding 40°C?

(b) If it is an intermittent load of 10 minutes duration followed by a 15 minutes shut down period, what is the maximum value of the load which the motor can take during the load period without temperature rise exceeding 40°C?

5.10 A motor equipped with a flywheel has to supply a load torque of 600 N · m for 10 sec. followed by a no-load period long enough for the flywheel to regain its full speed. It is desired to limit the motor torque to 450 N · m. The no-load speed of the motor is 600 rpm and it has 8% slip at a torque of 450 N · m. The speed-torque characteristic may be assumed to be a straight line in the region of interest. What should be the moment of inertia of the motor and flywheel?

CHAPTER 6

TRANSIENTS AND DYNAMICS

6.1 EQUATION OF MOTION

Whenever there is a transition of an electrical drive from one steady state condition to another, transients occur in the drive system during this period. During a transient period, which lasts for a short time, the speed, torque, current, etc. change from one steady state value to another. Electric motors undergo transient operation during starting, braking, speed control, speed reversal, etc. Transients may also be caused due to sudden changes in load, supply voltage and frequency.

The validity of motor capacity, the requisite power control equipment, the design of control circuits and the means of reducing the losses involved in starting, braking and speed reversal are based on the knowledge of the behaviour of electric drives under transient conditions.

A fundamental mechanical equation, known as the equation of motion, expresses the relationship between the driving torque developed by the motor and the load torque, based on Newton's law of motion:

$$J\frac{d\omega}{dt} - \frac{\omega^2}{2}\frac{dJ(\alpha)}{d\alpha} = T_m(t) - T_L(t) \tag{6.1}$$

where

T_m is the torque developed by the motor in N · m,

T_L is the torque developed by the mechanism in N · m,

ω is the angular speed of the motor shaft in rad/s,

J is the angular moment of inertia of the motor shaft in kg · m^2, and

$\frac{\omega^2}{2}\frac{dJ(\alpha)}{d\alpha}$ is the component of the equation which takes into account the variations in moment of inertia.

If the moment of inertia is constant, we have

$$\frac{dJ(\alpha)}{d\alpha} = 0$$

and the equation of motion can be rewritten in the following way:

$$J\frac{d\omega}{dt} = T_m(t) - T_L(t) \tag{6.2}$$

If the motor and the load torque are independent of time, Eq. (6.2) can be written as

$$J\frac{d\omega}{dt} = T_m - T_L \tag{6.3}$$

The fundamental mechanical relationships in dynamic as well as static conditions expressed by Eqs. (6.1–6.3) are now considered.

1. $T_m(t) > T_L(t)$, consequently $\frac{d\omega}{dt} > 0$ and therefore the drive experiences acceleration and the speed increases.

2. $T_m(t) < T_L(t)$, consequently $\frac{d\omega}{dt} < 0$ and therefore the drive experiences deceleration and the speed decreases.

3. $T_m(t) = T_L(t)$, consequently $\frac{d\omega}{dt} = 0$ and therefore the drive attains steady state, i.e. runs at constant speed.

The first two cases deal with transient processes (acceleration/deceleration) and Eqs. (6.1–6.3) are therefore called the *dynamic equations* of the drive.

The steady state or static equation of the drive is given by

$$T_m - T_L = 0$$

The torque T_m developed by the motor is sometimes called the ***active torque*** or ***driving torque***, while the torque T_L developed by the mechanism is called the *load torque* or ***resisting torque***.

6.2 EQUIVALENT SYSTEM

Generally, the different parts of a mechanism operate at different speeds. These parts are mechanically connected to the driving motor through gears, belts, etc. For transient analysis of the motor using the above mentioned dynamic equations, all mechanical quantities such as load torque, inertia torque, etc. should be referred to a single axis, i.e. motor shaft based on the principle of conservation of energy.

6.3 REFERRING TORQUE TO A SINGLE AXIS (MOTOR SHAFT)

Consider a general scheme of an electric drive shown in Fig. 6.1. It may be noticed that the motor develops the torque T_m on its shaft termed primary shaft, while the load torque T_L is applied to the output shaft of the gear train. To define their relationship, the law of conservation of energy is used.

Designating P'_L and P_L as powers on the primary, and output shafts, respectively, where P_L is the output power and P'_L referred to the primary shaft, the following relationship can be written.

$$P_L = \eta P'_L$$

where η is the overall efficiency of the gear train

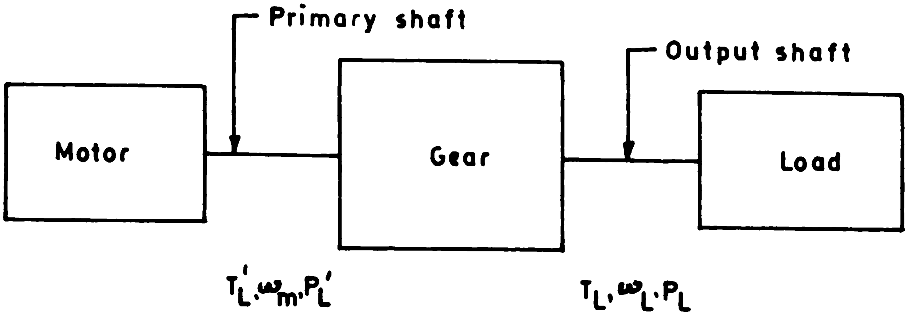


Fig. 6.1 Schematic diagram of a motor-load combination coupled through a gear train.

On the other hand, we have

$$P_L = T_L\omega_L \text{ and } P_L' = T_L'\omega_m$$

where T_L' is the load torque, and T_L referred to the output shaft. Then

$$T_L\omega_L = \eta T_L'\omega_m$$

or

$$T_L' = T_L\frac{\omega_L}{\eta\omega_m} = \frac{T_L}{\eta\left(\dfrac{\omega_m}{\omega_L}\right)} = \frac{T_L}{\eta\cdot i} \tag{6.4}$$

where i = gear ratio = $\dfrac{\omega_m}{\omega_L}$.

Considering that the gear train consists of several units with gear ratios $i_1, i_2, i_3, \ldots, i_n$, and the corresponding efficiencies are $\eta_1, \eta_2, \eta_3, \ldots, \eta_n$, Eq. (6.4) can be written as

$$T_L' = \frac{T_L}{(\eta_1\cdot\eta_2\cdot\eta_3\cdot\ldots\cdot\eta_n)(i_1\cdot i_2\cdot i_3\cdot\ldots\cdot i_n)} \tag{6.5}$$

In the same manner, the motor torque T_m can be referred to the output shaft. In this case

$$T_m' = T_m(\eta_1\cdot\eta_2\cdot\eta_3\cdot\ldots\cdot\eta_n)(i_1\cdot i_2\cdot i_3\cdot\ldots\cdot i_n) \tag{6.6}$$

It is significant to note that if the gear train reduces the speed from the motor to the mechanism, then the referred load torque is decreased too.

6.4 REFERRING MOMENT OF INERTIA TO A SINGLE AXIS (MOTOR SHAFT)

In Eqs. (6.1–6.3), the moment of inertia is assumed to be referred to the primary axis. An equivalent moment of inertia as referred to a single axis can be defined as the one, which expresses the inertia properties of the kinetic system in terms of the angular speed of a single shaft. Usually, the motor shaft or the output shaft is selected for this purpose. This referring is also done on the basis of the law of conservation of energy.

Consider the typical scheme shown in Fig. 6.2. Here the kinetic energy of all rotating parts can be expressed as equal to

$$\frac{1}{2}J_m\omega_m^2 + \frac{1}{2}J_1\omega_1^2 + \frac{1}{2}J_L\omega_L^2 + \text{energy stored in } W_L$$

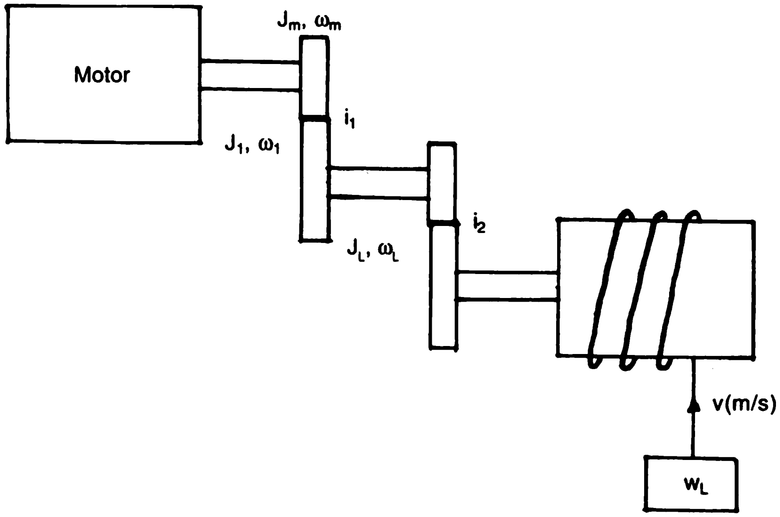


Fig. 6.2 A typical motor-load system with multiple gears.

Designating J as equivalent inertia referred to the primary shaft, the following equation can be written based on the principle of conservation of energy.

$$\frac{1}{2}J\omega_m^2 = \frac{1}{2}J_m\omega_m^2 + \frac{1}{2}J_1\omega_1^2 + \frac{1}{2}J_L\omega_L^2 \text{ (at no-load)}$$

or

$$J = J_m + J_1\left(\frac{\omega_1}{\omega_m}\right)^2 + J_L\left(\frac{\omega_L}{\omega_m}\right)^2$$

Designating $\frac{\omega_m}{\omega_1} = i_1$, $\frac{\omega_1}{\omega_L} = i_2$, we have

$$J = J_m + \frac{J_1}{(i_1)^2} + \frac{J_L}{(i_1 \cdot i_2)^2}$$

This equation shows that the moments of inertia of the parts other than that of the motor are decreased by the square of the gear ratios.

Extending this rule to a multiple gear train, a general equation for an overall equivalent moment of inertia can be derived as follows:

$$J = J_m + \frac{J_1}{(i_1)^2} + \frac{J_L}{(i_1 \cdot i_2)^2} + \ldots + \frac{J_n}{(i_1 \cdot i_2 \cdot i_3 \cdot \ldots \cdot i_n)^2} \tag{6.7}$$

It may be noted that the weight W_L has not been taken into account in the calculation of the equivalent moment of inertia.

6.5 REFERRING FORCES AND MASSES HAVING TRANSLATIONAL MOTION TO MOTOR SHAFT

If there are parts of a mechanism with a translational motion defined by the weight W_L and the translational speed v as in the case of a motor hoist system (Fig. 6.2), the overall moment of inertia is given by

$$J = J_m + \frac{J_1}{(i_1)^2} + \frac{J_L}{(i_1 \cdot i_2)^2} + \frac{W_L}{g}\left(\frac{v}{\omega_m}\right)^2 \tag{6.8}$$

The weight having translational motion is referred to the motor shaft on the basis of constant kinetic energy. That is

$$\frac{1}{2} \cdot \frac{W_L}{g}(v)^2 = \frac{1}{2} J'(\omega_m)^2$$

The moment of inertia of the weight with translational motion (Fig. 6.2) referred to the motor shaft is given by

$$J' = \frac{W_L}{g}\left(\frac{v}{\omega_m}\right)^2 \tag{6.9}$$

In the same manner, the equivalent moment of inertia referred to the output shaft of the gear train can be obtained and expressed in the following way:

$$J_0 = \left[J_m + \frac{W_L}{g}\left(\frac{v}{\omega_m}\right)^2\right](i_1 \cdot i_2 \cdot i_3 \cdot \ldots \cdot i_n)^2 + J_1(i_2 \cdot i_3 \cdot \ldots \cdot i_n)^2 + \ldots + J_n \tag{6.10}$$

6.6 MOMENT OF INERTIA

The relation for moment of inertia is given by

Moment of inertia = Mass × (radius of gyration or radius of inertia)^2

or

$$J = mp^2$$

where

J = moment of inertia,

m = mass, and

p = radius of inertia.

In SI units, m is in kg and J is in kg · m^2. If G is the weight,

$$\text{mass} = m = \frac{G}{g}$$

where g = acceleration due to gravity.

If torque $T = J\dfrac{d\omega}{dt}$ is expressed in N · m, J is in kg · m^2 or N · m · sec^2

then

$$1 \text{ newton} = \frac{1}{9.81} \text{ kgf}$$

6.6.1 Determination of Moment of Inertia

It is necessary to know the moment of inertia of the moving parts of an electrical drive for the study of transient performance. The moment of inertia can be determined either from the design data of the drive or from the data obtained by performing the following test.

Retardation test

At the beginning of the experiment, the motor is run at a speed slightly more than the normal. The power input to the motor is noted. Immediately after this, the supply to the motor is cut off. The motor continues to run due to stored energy in its rotating parts. But it gradually slows down and eventually stops as the stored energy is used up to meet the rotational losses.

An oscillogram of the motor speed during the retardation period is taken. Speed versus time, $N = f(t)$ is plotted (Fig. 6.3).

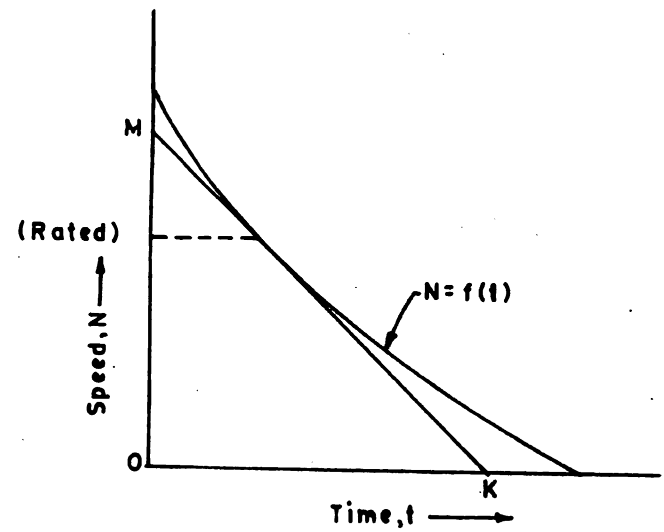


Fig. 6.3 Graphical method of determination of moment of inertia.

The power consumed to meet the rotational losses at any instant is given by

$$P = \frac{d}{dt}\left(\frac{1}{2}J\omega^2\right) = J\omega\frac{d\omega}{dt} = JN\frac{4\pi^2}{3600}\frac{dN}{dt} \tag{6.11}$$

where N is in rpm.

It may be noted that the rotational losses P just before switching off the power supply are equal to the rotational losses at the instant of switching off at rated speed. The inertia J can be calculated by using Eq. (6.10), and the data obtained experimentally.

At rated speed N, we have

$$\frac{dN}{dt} = \frac{OM}{OK}$$

6.7 OPTIMAL GEAR RATIO

The driving gear has radius r_1, angular speed ω_1, and angular acceleration ε_1. The driven gear has radius r_2, angular speed ω_2, and angular acceleration ε_2 (Fig. 6.4).

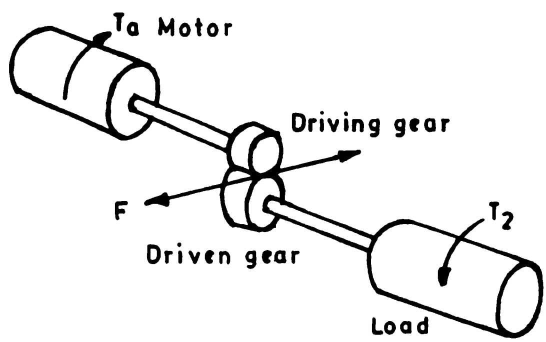


Fig. 6.4 A typical motor-load combination with gears.

Here,

$$\text{Gear ratio, } q = \frac{r_2}{r_1} = \frac{\omega_1}{\omega_2} = \frac{\varepsilon_1}{\varepsilon_2} \tag{6.12}$$

Let F be the tangential force between the gear teeth in contact. Motor torque T_a has to overcome the inertia torque $J_a\varepsilon_1$ to accelerate the rotor (armature) and the torque Fr_1 to drive the gear. Therefore

$$T_a = J_a\varepsilon_1 + Fr_1 \tag{6.13}$$

On the drive shaft, the torque Fr_2 has also to be equal to the inertia torque of the load $J_2\varepsilon_2$ and the load torque T_2. Therefore

$$Fr_2 = J_2\varepsilon_2 + T_2 \tag{6.14}$$

Now, Eq. (6.13) can be written as

$$T_a = J_a\varepsilon_1 + Fr_2\frac{r_1}{r_2} = J_a\varepsilon_1 + \frac{Fr_2}{q}$$

Using Eq. (6.14), we have

$$T_a = J_a\varepsilon_1 + \frac{J_2\varepsilon_2 + T_2}{q} \tag{6.15}$$

Now, from Eq. (6.12), $\varepsilon_1 = q\varepsilon_2$, therefore,

$$T_a = J_a q\varepsilon_2 + \frac{J_2\varepsilon_2 + T_2}{q}$$

Multiplying by q and arranging, we get

$$T_a q = J_a q^2\varepsilon_2 + J_2\varepsilon_2 + T_2$$

or

$$T_a = \left(J_a q + \frac{J_2}{q}\right)\varepsilon_2 + \frac{T_2}{q} \tag{6.16}$$

For optimum gear ratio, so that the motor torque T_a is minimum, we have

$$\frac{dT_a}{dq} = \frac{d}{dq}\left[\left(J_a q + \frac{J_2}{q}\right)\varepsilon_2 + \frac{T_2}{q}\right] = 0$$

or

$$\left(J_a - \frac{J_2}{q^2}\right)\varepsilon_2 - \frac{T_2}{q^2} = 0$$

or

$$\frac{J_2\varepsilon_2 + T_2}{q^2} = J_a\varepsilon_2$$

or

$$q = \sqrt{\frac{J_2\varepsilon_2 + T_2}{J_a\varepsilon_2}} \tag{6.17}$$

When the load torque T_2 is negligible, we get

$$q = \sqrt{\frac{J_2}{J_a}} \tag{6.18}$$

From Eq. (6.13), we already know that

$$T_a = J_a\varepsilon_1 + Fr_1$$

where

ε_1 = angular acceleration of the motor shaft = $\dfrac{d\omega_1}{dt}$, and

$Fr_1 = T_1$ is the torque required to drive the mechanism (at the motor shaft).

Then

$$J_a \varepsilon_1 = T_a - T_1$$

or

$$J_a \frac{d\omega_1}{dt} = T_a - T_1$$

T_a and T_1 are the functions of time. We can, therefore, write

$$J_a \frac{d\omega_1}{dt} = T_a(t) - T_1(t) \tag{6.19}$$

or

$$d\omega_1 = \frac{T_a(t) - T_1(t)}{J_a} dt$$

or

$$\omega_1 = \frac{1}{J_a} \int_{t_1}^{t_2} [T_a(t) - T_1(t)] dt \tag{6.20}$$

where

$\omega_1 = \omega_{t_2} - \omega_{t_1}$,

ω_{t_1} = angular speed at the instant t_1, and

ω_{t_2} = angular speed at the instant t_2.

If $\omega_{t_1} = 0$ at time t_1, then $\omega_{t_2} = \omega_1$.

The expression for time can be derived as shown below.

From Eq. (6.19), we get

$$dt = J_a \frac{d\omega_1}{T_a(t) - T_1(t)}$$

or

$$t = J_a \int_{\omega_{t_1}}^{\omega_{t_2}} \frac{d\omega}{T_a(t) - T_1(t)} \tag{6.21}$$

$$\omega_1 = f(t)$$

If the torque difference is assumed constant, we have

$$\Delta T = T_a(t) - T_1(t) = aT = \text{constant}$$

Then

$$t = J_a \int_{\omega_{t_1}}^{\omega_{t_2}} \frac{d\omega_1}{aT} = \frac{J_a(\omega_{t_2} - \omega_{t_1})}{aT} \tag{6.22}$$

If the torque difference is a linear function of speed, such that

$$\Delta T = a - b\omega_1$$

Then

$$t = J_a \int_{\omega_{t_1}}^{\omega_{t_2}} \frac{d\omega_1}{a - b\omega_1} = J_a b \log_e \left(\frac{a - b\omega_{t_1}}{a - b\omega_{t_2}} \right) \tag{6.23}$$

To determine the angular movement ϕ_1 of the motor shaft, we have

$$d\phi_1 = \omega_1 dt$$

But it can also be written that

$$\frac{d\omega_1}{dt} = \frac{d\omega_1}{d\phi_1} \cdot \frac{d\phi_1}{dt} = \omega_1 \frac{d\omega_1}{d\phi_1}, \text{ because } \omega_1 = \frac{d\phi_1}{dt}$$

Referring to Eq. (6.19), we can write

$$J_a \frac{d\omega_1}{dt} = J_a \omega_1 \frac{d\omega_1}{d\phi_1} = \Delta T$$

or

$$d\phi_1 = J_a \frac{\omega_1 d\omega_1}{\Delta T}$$

or

$$\phi_1 = J_a \int_{\omega_{t_1}}^{\omega_{t_2}} \frac{\omega_1 d\omega_1}{\Delta T} \tag{6.24a}$$

where

$\phi_1 = \phi_{t_2} - \phi_{t_1}$,

ϕ_{t_1} = angular position at the instant t_1, when $\omega_1 = \omega_{t_1}$, and

ϕ_{t_2} = angular position at the instant t_2, when $\omega_1 = \omega_{t_2}$.

In case, where ΔT is constant, we get

$$\phi_1 = \frac{1}{2} J_a \Delta T \left(\omega_{t_2}^2 - \omega_{t_1}^2 \right) \tag{6.24b}$$

If the torque difference is a linear function of speed, we can write

$$\Delta T = a - b\omega_1$$

Then

$$\phi_1 = J_a \int_{\omega_{t_1}}^{\omega_{t_2}} \frac{\omega_1 d\omega_1}{a - b\omega_1} = J_a \left[\int_{\omega_{t_1}}^{\omega_{t_2}} \frac{a d\omega_1}{b(a - b\omega_1)} - \int_{\omega_{t_1}}^{\omega_{t_2}} \frac{d\omega_1}{b} \right]$$

$$= J_a \left[a \log_e \left(\frac{a - b\omega_{t_1}}{a - b\omega_{t_2}} \right) + \frac{\omega_{t_1} - \omega_{t_2}}{b} \right] \qquad (6.25)$$

6.8 REFERRING TORQUES AND MASSES HAVING TRANSLATIONAL MOTION AT VARIABLE SPEEDS

Let us consider a crank gear of radius r connected to a translating mass (Fig. 6.5), which undergoes to and fro motion at variable speeds and acceleration during each revolution of the crankshaft.

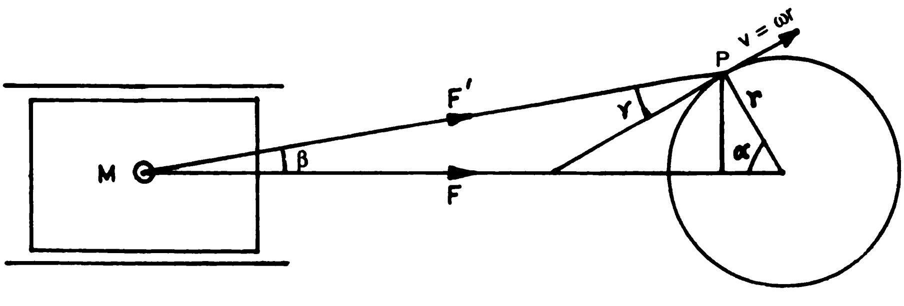


Fig. 6.5 A typical crank gear system.

The velocity v and hence the kinetic energy stored in the translating mass varies between zero and maximum.

The equivalent inertia of the translating mass referred to the crankshaft is

$$J = \frac{mv^2}{\omega^2} \qquad (6.26)$$

Referring to Fig. 6.5, we have

$$90° - \alpha = \beta + \gamma$$

or

$$\gamma = 90° - \alpha - \beta$$

Now

$$v \cos \beta = \omega r \cos (90° - \alpha - \beta) = \omega r \sin (\alpha + \beta)$$

or

$$v = \frac{\omega r \sin(\alpha + \beta)}{\cos \beta} \tag{6.27}$$

and

$$\beta = \sin^{-1}\left(\frac{r \sin \alpha}{l}\right) \tag{6.28}$$

β is, therefore, a function of α. Substituting v from Eq. (6.27) into Eq. (6.26), we get

$$J = \frac{mr^2(\sin(\alpha + \beta))^2}{(\cos \beta)^2} \tag{6.29}$$

The moment of inertia is also a function of angle α, i.e. depends upon the position of the crank-pin.

The total moment of inertia is the sum of the moment of inertia obtained from Eq. (6.29) and the moment of inertia of all the moving parts referred to the crankshaft. If there is some intermediate state of transmission with a gear train, the moment of inertia referred to the motor shaft may be determined by using Eq. (6.7). Let

F = Resisting force offered by the translating mass

i = Gear ratio, $\left(\frac{\omega_m}{\omega}\right)$

η = Gear efficiency

Referring to Fig. 6.5, we have

$$F' \cos \beta = F$$

or

$$F' = \frac{F}{\cos \beta}$$

The tangential force at the point P can be expressed as

$$F' \cos \gamma = F' \sin(\alpha + \beta) = \frac{F \sin(\alpha + \beta)}{\cos \beta}$$

The load torque T_L referred to the motor shaft is given by

$$T_L = \frac{Fr \sin(\alpha + \beta)}{\eta i \cos \beta} \tag{6.30}$$

The moment of inertia J is a variable quantity, being a function of angle α.

At any instant, the kinetic energy (KE) stored in the crankshaft is

$$\text{KE} = \frac{1}{2} J\omega^2$$

The corresponding dynamic power is given by

$$P_{dyn} = \frac{d}{dt}(\text{KE}) = \frac{d}{dt}\left(\frac{1}{2}J\omega^2\right) = J\omega\frac{d\omega}{dt} + \frac{1}{2}\omega^2\frac{dJ}{d\alpha}\frac{d\alpha}{dt} = J\omega\frac{d\omega}{dt} + \frac{1}{2}\omega^3\frac{dJ}{d\alpha} \tag{6.31}$$

The inertia torque, therefore, is

$$T_{dyn} = J\frac{d\omega}{dt} + \frac{1}{2}\omega^2\frac{dJ}{d\alpha} \tag{6.32}$$

If T_m is the torque developed by the motor, the equation of motion is given by

$$T_m - T_L = J\frac{d\omega}{dt} + \frac{1}{2}\omega^2\frac{dJ}{d\alpha}$$

The additional term on the right hand side is due to the variable moment of inertia.

6.9 TRANSIENTS IN DC MOTORS

DC motors are extensively used in industry, when precise control of speed and torque over a wide range in either direction is required. The common method is the adjustable voltage control of separately excited dc motors. The study of the transient characteristics during starting, braking and reversal becomes essential for designing the controller.

The dynamic behaviour of separately excited or shunt dc motors can be studied with the aid of the transfer function and block diagram representation. The transfer functions of interest relate speed and armature current to inputs as the changes in (a) terminal voltage, and (b) load torque. Magnetic saturation is neglected for simplicity. The field mmf acts along the direct axis and the armature mmf along the quadrature axis. There is no mutual inductance between the two windings. The demagnetizing effect of armature reaction is neglected. The study has been made for two cases: (a) when the armature inductance is negligibly small, and (b) when it is large enough to be considered.

The motor is started by closing the switch S (Fig. 6.6a). It is also assumed that the terminal voltage, the load torque and the field excitation remain unchanged. The equivalent inertia and the coefficient of viscous friction of the motor and the load are denoted by J and F, respectively. The basic equations are

$$V_t = iR_a + C\omega_m \tag{6.33}$$

$$iR_a = V_t - C\omega_m \tag{6.34}$$

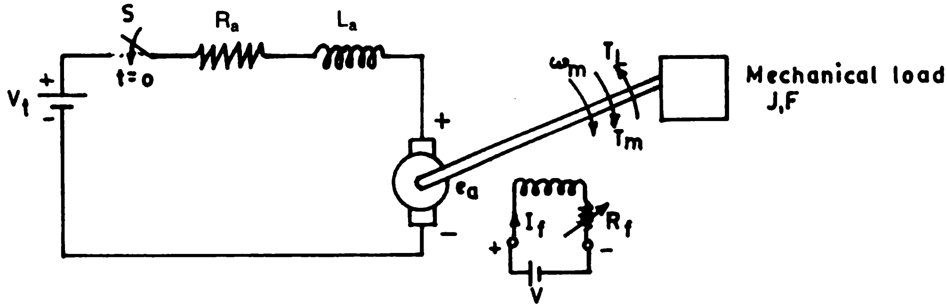
$$T = Ci = J\frac{d\omega_m}{dt} + F\omega_m + T_L \tag{6.35}$$

and

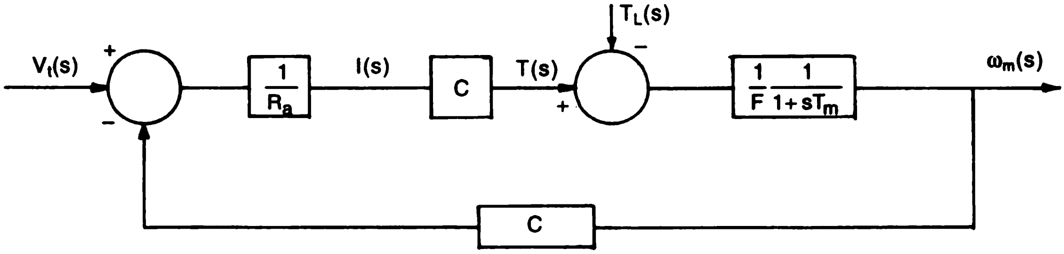
$$T - T_L = J\frac{d\omega_m}{dt} + F\omega_m \tag{6.36}$$

Laplace transformation of Eqs. (6.34) and (6.36) gives

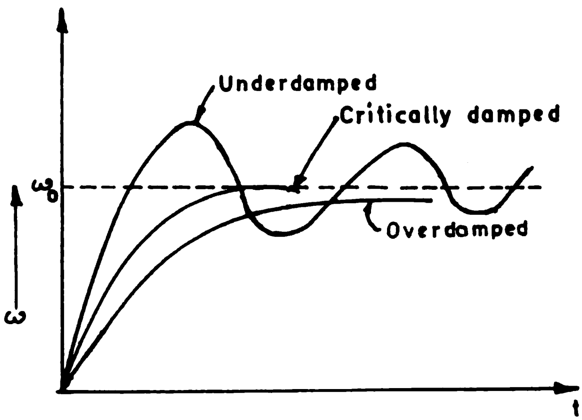
$$I(s)R_a = V_t(s) - C\Omega_m(s)$$



(a) Schematic diagram.



(b) Block diagram neglecting armature inductance.



(c) Dynamic response to step change in supply voltage.

Fig. 6.6 Separately excited dc motor with load.

$$T(s) - T_L(s) = Js\Omega_m(s) + F\Omega_m(s) \tag{6.37}$$

or

$$\frac{\Omega_m(s)}{T(s) - T_L(s)} = \frac{1}{Js + F} = \frac{1}{F(1 + sT_m)} \tag{6.38}$$

where T_m = mechanical time constant $= \dfrac{J}{F}$.

The block diagram representation of Eqs. (6.37) and (6.38), i.e. of a separately excited dc motor is shown in Fig. 6.6b.

By applying the feedback formula, the closed loop transfer function can be derived as

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{G(s)}{1+G(s)H(s)} = \frac{\dfrac{C}{R_aF}\cdot\dfrac{1}{(1+sT_m)}}{1+\dfrac{C^2}{R_aF}\cdot\dfrac{1}{(1+sT_m)}} = \frac{C}{(R_aF+C^2)}\cdot\frac{1}{(1+sT_{em})} \tag{6.39}$$

where T_{em} = Electromechanical time constant $= \dfrac{JR_a}{R_aF+C^2}$

Assuming the step change in terminal voltage to be V_t, the Laplace transform of $V_t(s) = \dfrac{V_t}{s}$ is substituted in Eq. (6.39) and it may then be shown that

$$\Omega_m(s) = \frac{\left(\dfrac{CV_t}{JR_a}\right)}{s\left(s+\dfrac{1}{T_{em}}\right)} = \frac{K_0}{s} + \frac{K_1}{s+\dfrac{1}{T_{em}}}$$

where

$$K_0 = \frac{\left(\dfrac{CV_t}{JR_a}\right)}{\left(s+\dfrac{1}{T_{em}}\right)} \text{ (at } s = 0) = \frac{CV_t}{R_aF+C^2}$$

and

$$K_1 = \frac{\left(\dfrac{CV_t}{JR_a}\right)}{s}\left(\text{at } s = -\frac{1}{T_{em}}\right) = -\frac{CV_t}{R_aF+C^2}$$

Therefore

$$\Omega_m(s) = \left(\frac{CV_t}{R_aF+C^2}\right)\left(\frac{1}{s} - \frac{1}{s+\dfrac{1}{T_{em}}}\right)$$

The corresponding time solution is

$$\omega_m(t) = \left(\frac{CV_t}{R_aF+C^2}\right)\left(1-e^{-\frac{t}{T_{em}}}\right) \tag{6.40}$$

If F is assumed to be zero, we get

$$\omega_m(t) = \frac{V_t}{C}\left(1 - e^{-\frac{t}{T_{em}}}\right) = \omega_0\left(1 - e^{-\frac{t}{T_{em}}}\right) \tag{6.41}$$

For the particular case, when the motor starts under load, we get

$$\omega_m(t) = \omega_L\left(1 - e^{-\frac{t}{T_m}}\right) \tag{6.42}$$

It shows that though the time required for reaching the final steady state speed is infinity, the transient process is practically completed within three to four times T_{em} (Fig. 6.6c).

The dynamic response of the motor to the step change in load torque is obtained by considering the transfer function between the speed and the load torque of the same block diagram (Fig. 6.6b). It may be noted that there is a transmission function of

$$\frac{1}{F(1 + sT_m)}$$

between $T_L(s)$ and $\Omega_m(s)$, whereas the corresponding feedback function is $\dfrac{C^2}{R_a}$.

The closed loop transfer function in this case is given by

$$\frac{\Omega_m(s)}{T_L(s)} = \frac{-\dfrac{1}{F(1+sT_m)}}{1 + \dfrac{C^2}{R_aF}\cdot\dfrac{1}{(1+sT_m)}} = -\frac{R_a}{(R_aF + C^2)}\cdot\frac{1}{(1+sT_{em})} \tag{6.43}$$

Assuming step change in load torque, we get

$$\Omega_m(s) = \frac{-T_LR_a}{(R_aF + C^2)}\cdot\frac{1}{s(1+sT_{em})}$$

Following the same procedure as employed for a step change in voltage V_t, the time expression for speed is obtained as under

$$\omega_m(t) = \frac{-T_LR_a}{(R_aF + C^2)}\left(1 - e^{-\frac{t}{T_{em}}}\right) \tag{6.44}$$

6.10 TRANSIENTS IN DC MOTORS WITH ARMATURE INDUCTANCE

The study of electrical transients in the armature circuit and mechanical transients in the mechanical system driven by the separately excited dc motor has been considered here, when the effect of armature inductance is not negligible.

The first step in obtaining the speed transfer function is to apply Kirchhoff's voltage law for the armature circuit. Magnetic linearity is assumed. The basic equation is

$$V_t = iR_a + L_a\frac{di_a}{dt} + C\omega_m \tag{6.45}$$

After Laplace transform, we get

$$I(s)R_a(1 + sT_a) = V_t(s) - C\Omega_m(s) \tag{6.46}$$

where electrical (armature) time constant $T_a = \dfrac{L_a}{R_a}$, or

$$\frac{I(s)}{V_t(s) - C\Omega_m(s)} = \frac{1}{R_a(1+sT_a)} \tag{6.47}$$

Hence, the block diagram of the system, when the armature inductance is taken into account, is modified as shown in Fig. 6.7. Using standard techniques, the block diagram is simplified as shown in Fig. 6.8. The load torque T_L has been converted to equivalent input voltage.

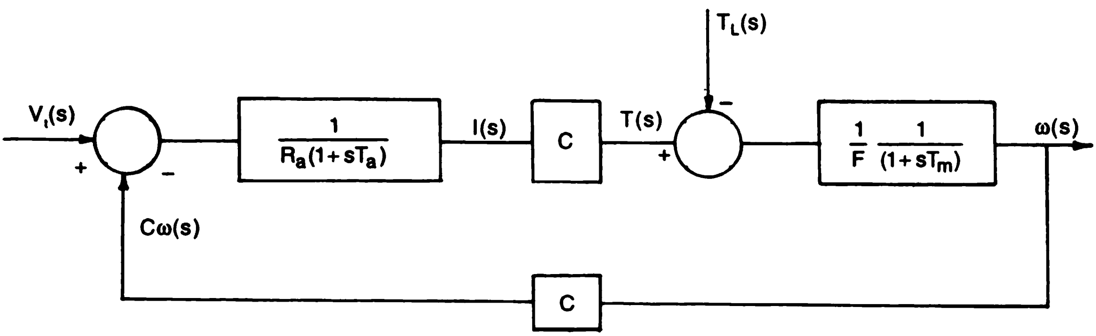


Fig. 6.7 Block diagram of the system of Fig. 6.6a with armature inductance.

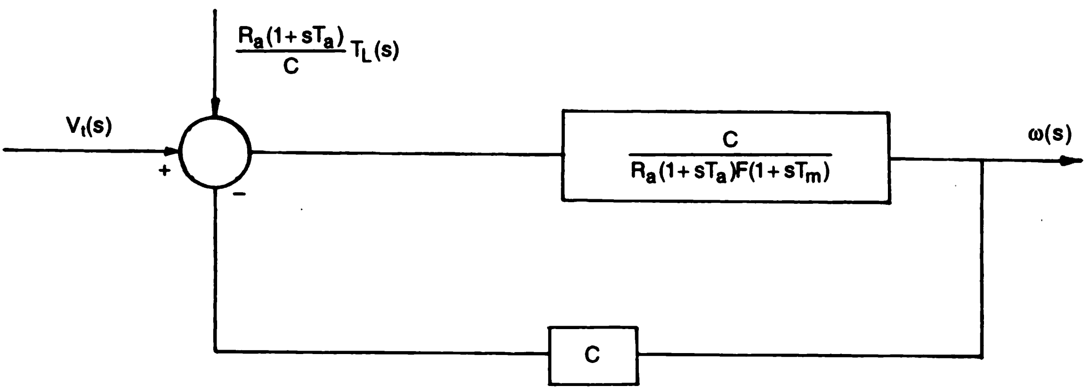


Fig. 6.8 Simplified block diagram of Fig. 6.7 for speed-voltage transfer function.

The speed/voltage transfer function is given as

$$\frac{\Omega_m(s)}{V_t(s) - \dfrac{R_a(1+sT_a)}{C}T_L(s)} = \frac{\dfrac{C_1}{A_1}}{1+\dfrac{B_1}{A_1}s+\dfrac{C_1}{A_1}s^2} \tag{6.48}$$

where $A_1 = R_aF + C^2$, $B_1 = R_aJ + L_aF$, $C_1 = L_aJ$.

The system is, therefore, of second order.

By application of the feedback formula it is found that the motor speed is related to the change in terminal voltage as

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{\dfrac{C}{R_aF(1+sT_a)(1+sT_m)}}{1+\dfrac{C^2}{R_aF(1+sT_a)(1+sT_m)}}$$

The above equation is simplified to

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{\dfrac{C}{R_aFT_aT_m}}{s^2+\left(\dfrac{T_a+T_m}{T_aT_m}\right)s+\left(\dfrac{1+\dfrac{C^2}{R_aF}}{T_aT_m}\right)} \tag{6.49}$$

The response according to Eq. (6.49) is of second order, because of two time constants, T_m and T_a, and also because of the values of C, J, F, L_a and R_a. The response may become underdamped or overdamped.

The characteristic equation of the system is

$$s^2+\left(\frac{T_a+T_m}{T_aT_m}\right)s+\left(\frac{1+\dfrac{C^2}{R_aF}}{T_aT_m}\right) = 0 \tag{6.50}$$

The above equation is of the general form

$$s^2 + 2\zeta\omega_n s + \omega_n^2 = 0$$

and is related to the physical form

$$J\frac{d^2\theta_0}{dt^2} + F\frac{d\theta_0}{dt} + K\theta_0 \tag{6.51}$$

where J is the total system moment of inertia referred to the output shaft, F is the total system viscous friction coefficient and K is the total system stiffness measured at the output shaft. The knowledge of these three coefficients completely specifies the system behaviour.

Example 6.1

A separately excited dc motor generates an open circuit voltage of 220 V at 1500 rpm and with a field current of 0.5 A. The machine parameters are:

$$R_a = 1\ \Omega, \qquad L_a = 0, \qquad F = 0$$

The combined inertia of the motor with load is 2.0 kg · m². The motor drives a constant load torque $T_L = 20$ N · m. The armature terminals are suddenly connected to 220 V dc supply, when the field current I_f is 0.5 A.

(a) Derive the expressions for the speed ω_m and the armature current i as a function of time.

(b) Determine the steady-state values of speed and armature current.

Solution

$V_t = 220$ V, $\quad N_m = 1500$ rpm, $\quad R_a = 1\ \Omega$, $\quad T_L = 20$ N · m, $\quad J = 2$ kg · m²,

$$\omega_m = \frac{2\pi}{60} N_m = \frac{2\pi}{60} \times 1500 = 157.1 \text{ rad/s}$$

(a) $V_t \approx E_b = C\omega_m$ or $C = \dfrac{E_b}{\omega_m} = \dfrac{220}{157.1} = 1.4$ V · s/rad

Using Eqs. (6.33) and (6.35) and substituting $F = 0$ in Eq. (6.35), we get

$$V_t = C\omega_m + \frac{JR_a}{C}\frac{d\omega_m}{1.4} + \frac{T_L R_a}{C}$$

or

$$220 = 1.4\omega_m + \frac{1.0 \times 2.0}{1.4}\frac{d\omega_m}{dt} + \frac{1.0 \times 20.0}{1.4} = 1.4\omega_m + 1.4\frac{d\omega_m}{dt} + 14.0$$

Taking the Laplace transforms, we get

$$\frac{220}{s} = 1.4\Omega_m(s) + 1.4s\Omega_m(s) + \frac{14}{s} = 1.4(s+1)\Omega_m(s) + \frac{14}{s}$$

or

$$\Omega_m(s) = \frac{220 - 14}{1.4s(s+1)} = \frac{147}{s(s+1)} = \frac{K_0}{s} + \frac{K_1}{s+1} = 147\left(\frac{1}{s} - \frac{1}{s+1}\right)$$

Taking the inverse Laplace transform, we get

$$\omega_m(t) = 147(1 - e^{-t}) \text{ rad/s}$$

Now

$$i(t) = \frac{V_t - C\omega_m(t)}{R_a} = \frac{220 - 1.4\omega_m(t)}{1.0} = 220 - 1.4 \times 147(1 - e^{-t}) = (14 + 206e^{-t}) \text{ A}$$

(b) Steady state speed, $\omega_m(\infty) = 147$ rad/s

Steady state current, $I(\infty) = 14$ A

Example 6.2

Derive the transfer function of the system shown in Fig. 6.6a, if the load torque is proportional to the speed.

Solution

$T_L \propto \omega_m$ or $T_L = F_L\omega_m$

Let the inertia of the system including the load be J.

Eq. (6.35) can be written as

$$T = Ci = J\frac{d\omega_m}{dt} + F\omega_m + F_L\omega_m$$

Substituting $B = F + F_L$, we get

$$Ci = J\frac{d\omega_m}{dt} + B\omega_m$$

or

$$CI(s) = Js\Omega_m(s) + B\Omega_m(s)$$

The mechanical time constant of the system under this condition is

$$T_m = \frac{J}{B} = \frac{J}{F + F_L}$$

Using Eq. (6.46) and the transformed equation for torque, we get

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{1}{C + \frac{BR_a}{C}(1 + sT_a)(1 + sT_m)} \tag{6.52}$$

Example 6.3

Derive the transfer function of the system shown in Fig. 6.6a, when the supply is disconnected.

Solution

The supply is suddenly disconnected by opening the switch S at $t = 0$ in Fig. 6.6a.

The dynamic equation of the mechanical system is

$$T = Ci = J\frac{d\omega_m}{dt} + F\omega_m = 0 \tag{6.53}$$

or

$$F\omega_m = -J\frac{d\omega_m}{dt} \tag{6.54}$$

The Laplace transform is

$$F\Omega_m(s) = -J(s\Omega_m(s) - \omega_0)$$

where ω_0 is the initial speed. Hence

$$\Omega_m(s) = \frac{J\omega_0}{F + Js} = \frac{J}{\left(s + \frac{1}{T_m}\right)}$$

Therefore, the time domain response of the speed is

$$\omega_m(t) = \omega_0 e^{-\frac{t}{T_m}} \tag{6.55}$$

Example 6.4

A load is being hauled in the fashion shown in Fig. 6.9. Determine the dynamics of the system with the motor speed ω_m as the output and the motor armature voltage V_a as the input. Assume the weight of the load W_L = constant, drum diameter = d, and the gear inertia = J_G.

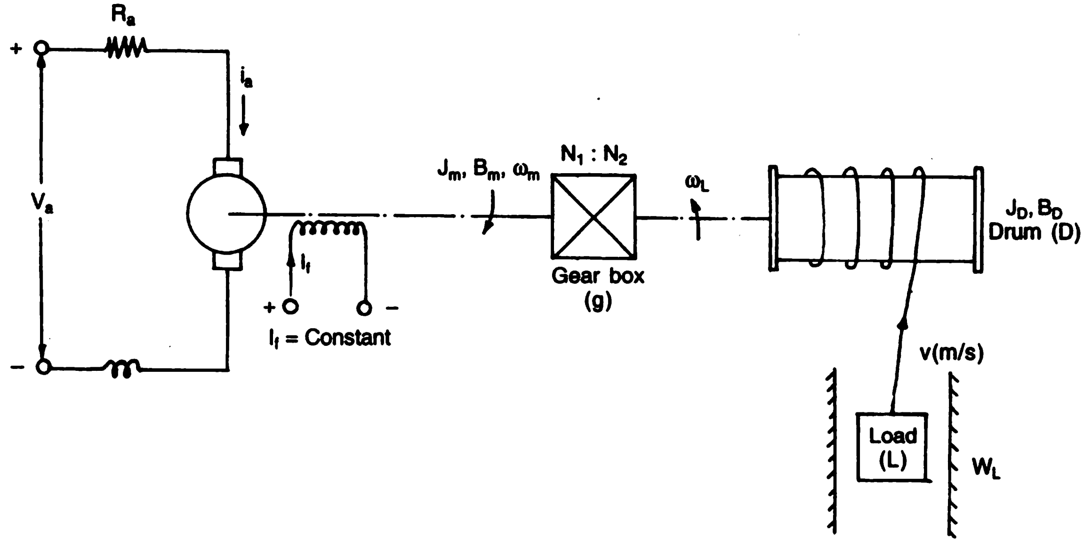


Fig. 6.9 Schematic diagram of the system of Example 6.4.

Solution

The mechanical quantities have to be referred to the motor shaft. Let us denote all referred quantities as shown in the above figure. Let the load be hauled at a velocity v (m/sec). The weight of the load is W_L, whereas the drum is rotating at a speed ω_L such that

$$\frac{\omega_L}{\omega_m} = \frac{N_1}{N_2}$$

(a) The drum inertia is to be referred to the motor shaft. Hence

$$\frac{1}{2}J_D(\omega_L)^2 = \frac{1}{2}J'_D(\omega_m)^2$$

or

$$J'_D = \left(\frac{\omega_L}{\omega_m}\right)^2 J_D = \left(\frac{N_1}{N_2}\right)^2 J_D$$

(b) Hoist load

The load torque is constant and equal to $T_{Lh} = \frac{1}{2}dW_L$. Therefore

$$T'_{Lh} = \frac{\omega_L}{\omega_m}T_{Lh} = \frac{N_1}{N_2}T_{Lh}$$

There is also the inertia of J_{Lh} corresponding to the kinetic energy stored by the weight W_L Therefore

$$\frac{1}{2}J_{Lh}\omega_L^2 = \frac{W_L}{2g}v^2$$

or

$$J_{Lh} = \frac{W_L}{g}\left(\frac{v}{\omega_L}\right)^2$$

It may be noted that $v = \frac{1}{2}d\omega_L$.

J_{Lh} has to be referred to the motor shaft. Therefore

$$J'_{Lh} = \left(\frac{\omega_L}{\omega_m}\right)^2 J_{Lh} = \left(\frac{N_1}{N_2}\right)^2 J_{Lh}$$

(c) Gear inertia

$$J'_G = \frac{J_G}{i^2} \qquad \text{where} \qquad i = \frac{N_1}{N_2}$$

(d) The viscous friction causes a load torque proportional to the speed. That is

$$T_v \propto \omega_L \qquad \text{or} \qquad T_v = K_1\omega_L$$

Also

$$T'_v = \frac{\omega_L}{\omega_m}T_v = \frac{\omega_L}{\omega_m}K_1\omega_L$$

The total moment of inertia referred to the motor shaft is

$$J = J_m + \frac{J_G}{i^2} + J'_D + J'_{Lh}$$

The total torque referred to the motor shaft is

$$T_L = T'_{Lh} + T'_v$$

The equations of the dc motor are same as Eqs. (6.45) and (6.35). The output-input relation in terms of transfer functions, i.e. in frequency domain, is given by

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{1}{C + \frac{FR_a}{C}(1 + sT_a)(1 + sT_m)} \tag{6.56}$$

which is a second order system, being same as Eq. (6.52).

The block diagram of the system is shown below in Fig. 6.10.

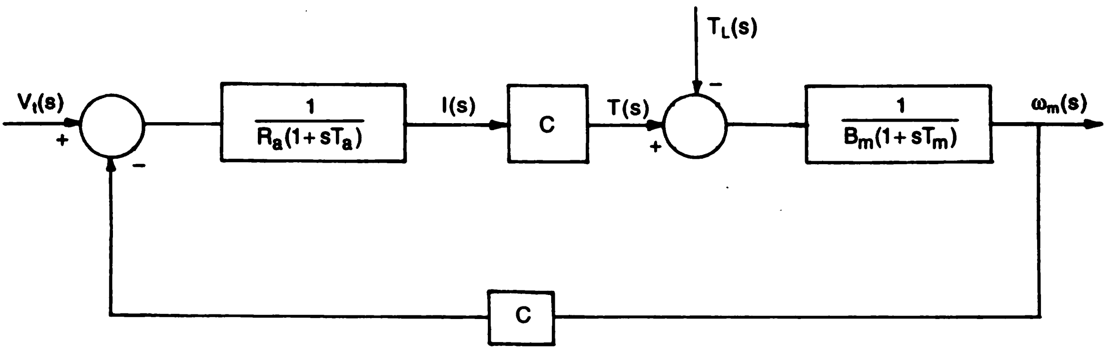


Fig. 6.10 Block diagram of the system of Fig. 6.9.

6.11 DYNAMIC BRAKING OF DC SHUNT MOTORS

Whenever the supply to a running shunt motor is disconnected and a resistance shunted across its armature, the motor immediately develops braking torque (Fig. 2.6). The field excitation remains unaltered. The current in the armature is reversed and braking torque is produced.

The equation for armature circuit during dynamic braking is

$$C\omega_m + iR = 0 \tag{6.57}$$

where $R = R_a + R_{db}$.

The dynamic equation of the mechanical system is

$$T = Ci = J\frac{d\omega_m}{dt} + F\omega_m + T_L$$

or

$$i = \frac{J}{C}\frac{d\omega_m}{dt} + \frac{F}{C}\omega_m + \frac{T_L}{C} \tag{6.58}$$

Substituting Eq. (6.58) in Eq. (6.57), we get

$$C\omega_m + \frac{RJ}{C}\frac{d\omega_m}{dt} + \frac{RF}{C}\omega_m + \frac{R}{C}T_L = 0$$

or

$$T_{em}\frac{d\omega_m}{dt} + \omega_m + \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L = 0$$

where $T_{em} = \dfrac{JR}{RF + C^2}$, $\quad \Delta\omega_L = \dfrac{RT_L}{C^2}$

Taking Laplace transforms, we have

$$T_{em}(s\Omega_m(s) - \omega_{in}) + \Omega_m(s) + \left(\frac{C^2}{RF + C^2}\right)\frac{\Delta\omega_L}{s} = 0$$

or

$$\Omega_m(s) = \frac{\omega_{in}}{s + \frac{1}{T_{em}}} - \frac{C^2}{(RF + C^2)T_{em}}\frac{\Delta\omega_L}{s\left(s + \frac{1}{T_{em}}\right)}$$

$$= \frac{\omega_{in}}{s + \frac{1}{T_{em}}} - \frac{K_0}{s} - \frac{K_1}{s + \frac{1}{T_{em}}}$$

where $K_0 = \left(\dfrac{C^2}{RF + C^2}\right)\Delta\omega_L$, $\quad K_1 = -\left(\dfrac{C^2}{RF + C^2}\right)\Delta\omega_L$

Therefore

$$\Omega_m(s) = \frac{\omega_{in}}{s + \frac{1}{T_{em}}} - \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L\left[\frac{1}{s} - \frac{1}{s + \frac{1}{T_{em}}}\right]$$

Taking the inverse Laplace transform, we get

$$\omega_m(t) = -\left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L + \left[\omega_{in} + \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L\right]e^{-\frac{t}{T_{em}}} \tag{6.59}$$

6.11.1 Time of Braking

Supposing the motor is braked to speed ω_1 in time t_1 from initial speed ω_{in}, then

$$\omega_1 = -\left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L + \left[\omega_{in} + \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L\right]e^{-\frac{t_1}{T_{em}}}$$

$$t_1 = \text{time of braking} = T_{em} \log_e\left[\frac{\omega_{in} + \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L}{\omega_1 + \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L}\right] \tag{6.60}$$

6.11.2 Current during Dynamic Braking

During dynamic braking, the equations are

$$C\omega_m + iR = 0$$

and

$$i = -\frac{C\omega_m}{R}$$

Substituting for ω_m from Eq. (6.59), we get

$$i = \frac{C^2}{RF + C^2}\frac{C}{R}\Delta\omega_L - \frac{C}{R}\left[\omega_{in} + \left(\frac{C^2}{RF + C^2}\right)\Delta\omega_L\right]e^{-\frac{t}{T_{em}}}$$

$$= \frac{C^2}{RF + C^2}I_L - \left[I_{in} + \left(\frac{C^2}{RF + C^2}\right)I_L\right]e^{-\frac{t}{T_{em}}} \tag{6.61}$$

where

$$I_L = \frac{C}{R}\cdot\frac{T_L R}{C^2} = \frac{T_L}{C}$$

If the braking takes place at no load, we have $I_L = 0$. Then

$$i = -I_{in}e^{-\frac{t}{T_{em}}} \tag{6.62}$$

where

$$I_{in} = \frac{C\omega_{in}}{R} \tag{6.63}$$

If the friction is neglected

$$i = -(I_{in} + I_L)e^{-\frac{t}{T_{em}}} + I_L \tag{6.64}$$

The braking time can also be determined from the braking current using either Eq. (6.61) or Eq. (6.64) as the case may be. On the basis of Eq. (6.64), the time of braking is equal to

$$t_1 = T_{em} \log_e\left(\frac{I_{in} + I_L}{I_L}\right)$$

6.12 COUNTER CURRENT BRAKING AND SPEED REVERSAL OF DC SHUNT MOTORS

The counter current braking (plugging) or speed reversal of shunt motors is accomplished by interchanging the polarity of the voltage applied to the armature terminals of the motor as described in Section 2.2.1(b) (Fig. 2.7). The field excitation remains unaltered.

During counter current braking, the equations are

$$-V = C\omega_m + iR \tag{6.65}$$

$$T = Ci = J\frac{d\omega_m}{dt} + F\omega_m + T_L$$

or

$$i = \frac{J}{C}\frac{d\omega_m}{dt} + \frac{F}{C}\omega_m + \frac{T_L}{C} \tag{6.66}$$

It may be noticed that the Eqs. (6.33 and 6.35), (6.57 and 6.58), and (6.65 and 6.66) are basically the same equations with different armature voltages for starting, dynamic braking, and counter current braking, respectively.

Substituting for i from Eq. (6.66) in Eq. (6.65), we get

$$-V = C\omega_m + \frac{RJ}{C}\frac{d\omega_m}{dt} + \frac{RF}{C}\omega_m + \frac{R}{C}T_L$$

or

$$T_{em}\frac{d\omega_m}{dt} + \omega_m + \frac{C^2}{RF + C^2}(\Delta\omega_L + \omega_0) = 0$$

where $\omega_0 = \dfrac{V}{C}$.

Taking the Laplace transforms, we have

$$T_{em}(s\Omega_m(s) - \omega_m) + \Omega_m(s) + \left(\frac{C^2}{RF + C^2}\right)\frac{(\Delta\omega_L + \omega_0)}{s} = 0$$

or

$$\Omega_m(s) = \frac{\omega_{in}}{s + \frac{1}{T_{em}}} - \frac{C^2}{(RF + C^2)T_{em}} \frac{(\Delta\omega_L + \omega_0)}{s\left(s + \frac{1}{T_{em}}\right)}$$

$$= \frac{\omega_{in}}{s + \frac{1}{T_{em}}} - \frac{K_0}{s} - \frac{K_1}{s + \frac{1}{T_{em}}}$$

where

$$K_0 = \left(\frac{C^2}{RF + C^2}\right)(\Delta\omega_L + \omega_0), \qquad K_1 = -\left(\frac{C^2}{RF + C^2}\right)(\Delta\omega_L + \omega_0)$$

Therefore

$$\Omega_m(s) = \frac{\omega_{in}}{s + \frac{1}{T_{em}}} - \left(\frac{C^2}{RF + C^2}\right)(\Delta\omega_L + \omega_0)\left[\frac{1}{s} - \frac{1}{s + \frac{1}{T_{em}}}\right]$$

Taking the inverse transform, we get

$$\omega_m(t) = -\frac{C^2}{RF + C^2}(\Delta\omega_L + \omega_0) + \left[\omega_{in} + \left(\frac{C^2}{RF + C^2}\right)(\Delta\omega_L + \omega_0)\right]e^{-\frac{t}{T_{em}}} \tag{6.67}$$

When the motor is reversed without load, we have

$$\omega_{in} = \omega_L = \omega_0, \qquad \text{and} \qquad \Delta\omega_L = 0$$

F is assumed to be zero. Then

$$\omega_m(t) = -\omega_0 + 2\omega_0 e^{-\frac{t}{T_{em}}} \tag{6.68}$$

For derivation of the expression for current, we have

$$-V = C\omega_m + iR$$

or

$$i = -\frac{V + C\omega_m}{R}$$

It is assumed that $F = 0$, $\omega_{in} = \omega_L$, and $\omega_0 = \frac{V}{C}$. Then

$$i = -\frac{V}{R} + \frac{C}{R}(\omega_0 + \Delta\omega_L) - \left[\frac{C}{R}\Delta\omega_L + \frac{C}{R}(\omega_0 + \Delta\omega_L)\right]e^{-\frac{t}{T_{em}}}$$

Now

$$\frac{V}{R} = \frac{C}{R}\omega_0, \quad I_L = \frac{C}{R}\Delta\omega_L, \quad \text{and} \quad I_{in} = \frac{C}{R}(\omega_0 + \Delta\omega_L)$$

where R = total resistance in the armature circuit. Therefore

$$i = -(I_{in} + I_L)e^{-\frac{t}{T_{em}}} + I_L \tag{6.69}$$

If the reversal takes place without load, we have $I_L = 0$. Then

$$i = -I_{in}e^{-\frac{t}{T_{em}}}$$

where $I_{in} = \dfrac{2V}{R} = 2I_{sc}$. Hence

$$i = -\,2I_{sc}e^{-\frac{t}{T_{em}}} \tag{6.70}$$

Example 6.5

The following parameters are given for a separately excited dc motor:

240 V, 75 kW, 1750 rpm, $R_a = 0.015\ \Omega$, $L_a = 0.001$ H,
$C = 1.27$ V · s/rad, $J = 3.7$ kg · m^2, $F = 0$

Assume that the load inertia is equal to the motor inertia. Find the transfer function indicating the natural frequency and damping ratio.

Solution

$R_a = 0.015\ \Omega$, $L_a = 0.001$ H, $J = 3.7$ kg · m^2, $C = 1.27$ V · s/rad

Armature time constant,

$$T_a = \frac{L_a}{R_a} = \frac{0.001}{0.015} = 0.0667 \text{ sec.}$$

Electromechanical time constant,

$$T_{em} = \frac{JR_a}{C^2} = \frac{7.4 \times 0.015}{(1.27)^2} = 0.0688 \text{ sec.}$$

Friction F has been neglected.

The block diagram is shown in Fig. 6.11. Now

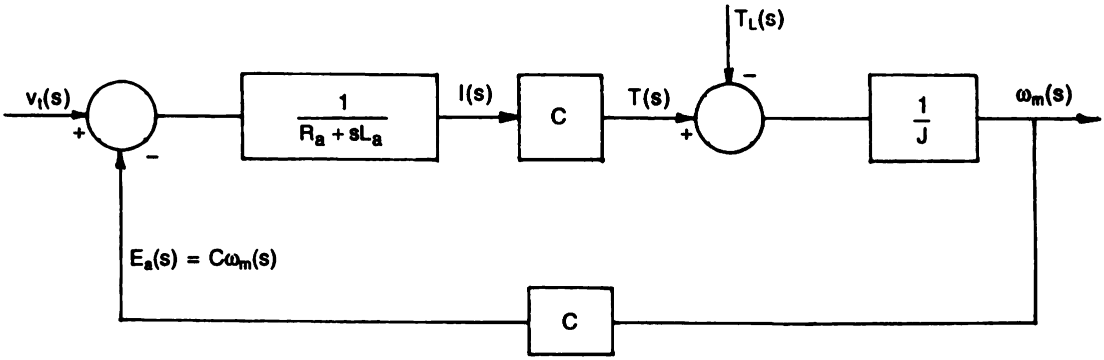


Fig. 6.11 Block diagram of the system of Example 6.5.

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{\dfrac{C}{Js(R_a + sL_a)}}{1 + \dfrac{C}{Js(R_a + sL_a)}} = \frac{1}{C} \cdot \frac{1}{1 + \left(\dfrac{JR_a}{C^2}\right)s + \left(\dfrac{JR_a}{C^2} \cdot \dfrac{L_a}{R_a}\right)s^2} = \frac{1}{C[1 + sT_{em}(1 + sT_a)]}$$

$$= \frac{1}{1.27[1 + 0.0688s(1 + 0.0667s)} = \frac{171.6}{s^2 + 15.0s + 218.0}$$

Alternatively, we can write the transfer function in the standard form of a second order system.

$$\frac{\Omega_m(s)}{V_t(s)} = \frac{k\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

where ω_n is the natural frequency of oscillation given by

$$\omega_n = \frac{1}{\sqrt{T_a T_{em}}}$$

and the damping ratio ζ is obtained as under:

$$2\zeta\omega_n = \frac{1}{T_a}$$

or

$$\zeta = \frac{1}{T_a} \cdot \frac{1}{2\omega_n} = \frac{1}{2}\sqrt{\frac{T_{em}}{T_a}}$$

Therefore

$$\omega_n = \sqrt{218} = 14.76 \text{ rad/s}$$

Now,

$$2\zeta\omega_n = 15.0$$

or

$$\zeta = \frac{15.0}{2 \times 14.76} = 0.5$$

Example 6.6

For the motor, the ratings of which are given in Example 6.5, determine the time equation of the change in speed for a constant input voltage and step change in load torque by 200 N · m.

Solution

Transfer function

$$\frac{\Omega_m(s)}{T_L(s)} = -\frac{R_a}{C^2} \cdot \frac{(1 + sT_a)}{1 + sT_{em} + (1 + sT_a)} = -\frac{0.015}{(1.27)^2} \cdot \frac{(1 + 0.0667s)}{1 + 0.0688s(1 + 0.0667s)}$$

$$= -\frac{2(1 + 0.0667s)}{s^2 + 15.0s + 218.0}$$

For the step change in T_L by 200 N · m, we have

$$\Omega_m(s) = -\frac{400(1 + 0.0667s)}{s(s^2 + 15.0s + 218.0)}$$

Then

$$\Omega_m(s) = -\frac{1.835}{s} + \frac{1.6 - j0.22}{s + 7.5 + j12.72} + \frac{1.6 + j0.22}{s + 7.5 - j12.72}$$

Taking the inverse Laplace transform, we have

$$\omega_m(t) = -1.835 + (3.2 \cos 12.72t - 0.44 \sin 12.72t)e^{-7.5t}$$

6.13 ENERGY ASSOCIATED WITH TRANSIENT PROCESS OF DC SHUNT MOTORS

The power drawn from the supply by a dc shunt motor to develop mechanical power output at the shaft is used up in

(a) carrying load at the shaft,

$$P_L = T_L \omega_m$$

(b) providing starting energy to the rotating parts of the drive system, and the dynamic torque determines the power required.

$$P_{dyn} = T_{dyn}\omega_m = J\omega_m = \frac{d\omega_m}{dt}$$

(c) supplying the losses in the motor.

Windage and friction losses are assumed to be included in the mechanical power output. The excitation loss is considered to be negligibly small. The armature copper loss is I^2R in the armature circuit.

6.13.1 Energy Loss during Starting

The energy loss is obtained from the equation

$$A_L = \int_0^{t_{st}} i^2 R dt \tag{6.71}$$

The KVL equation during starting of the motor in one step is

$$V = E + iR$$

or

$$i^2R = Vi - Ei = \omega_0 T - \omega T \tag{6.72}$$

Case I

The motor is started without load. Friction is neglected.

$$T = J\frac{d\omega_m}{dt}$$

or

$$dt = \frac{J}{T} d\omega_m \tag{6.73}$$

Substituting the quantities of Eqs. (6.72) and (6.73) in Eq. (6.71), we have

$$A_L = \int_{\omega_{in}}^{\omega_{fin}} J(\omega_0 - \omega_m) d\omega_m \tag{6.74}$$

If started from rest, $\omega_{in} = 0$ and $\omega_{fin} = \omega_0$. Then

$$A_L = \frac{1}{2} J\omega_0^2 \tag{6.75}$$

It shows that the energy loss in the motor during starting is equal to the stored energy in its rotating parts at steady state speed irrespective of the armature circuit resistance, the number of steps in the starting resistance, the resistance of each step and starting time.

The work done by the motor in storing kinetic energy in its rotating parts is given by

$$A_{mech} = \int_0^{\omega_0} J\omega_m \frac{d\omega_m}{dt} dt = \int_0^{\omega_0} J\omega_m d\omega_m = \frac{1}{2} J\omega_0^2 \tag{6.76}$$

Therefore, the amount of electrical energy drawn by the motor during starting is equal to double the kinetic energy stored in it. That is

$$A_{elec} = A_L + A_{mech} = J\omega_0^2 \tag{6.77}$$

Case II

The motor is started up with constant load torque. Therefore

$$i^2R = T(\omega_0 - \omega_m) = (T_L + T_j)(\omega_0 - \omega_m)$$

Now,

$$T_j = J\frac{d\omega_m}{dt}$$

Thus, energy loss during starting is

$$A_L = \int_0^{t_{st}} i^2R\,dt = \int_0^{\omega_L} J(\omega_0 - \omega_m)\,d\omega_m + \int_0^{t_{st}} T_L(\omega_0 - \omega_m)dt$$

$$= J\left[\omega_0\omega_L - \frac{1}{2}\omega_L^2\right] + T_L\left[\omega_0 t_{st} - \int_0^{t_{st}} \omega_m dt\right] \tag{6.78}$$

The first part of Eq. (6.78) represents the energy loss in the armature circuit due to acceleration, whereas the second part represents the loss in the armature circuit on account of load carried by the motor.

6.13.2 Energy Loss during Dynamic Braking

Case I

If dynamic braking takes place without load ($T_L = 0$), then $V = 0$, $\omega_{in} = \omega_0$, and $\omega_{fin} = 0$.
Equation (6.74) may be referred to for this case. The energy loss is

$$A_{db} = \int_{\omega_{in}}^{\omega_{fin}} -J\omega_m d\omega_m = -J\int_{\omega_0}^{0} \omega_m d\omega_m = \frac{1}{2}J\omega_0^2 \tag{6.79}$$

So, the energy loss during dynamic braking is equal to the stored energy in the rotating parts of the machine, when braking is applied.

Case II

If dynamic braking is applied with load torque, based on Eq. (6.78), the energy loss is

$$A_{db} = \int_{\omega_L}^{0} J(-\omega_m)d\omega_m + \int_0^{t_b} T_L(-\omega_m)dt = \frac{1}{2}J\omega_L^2 - T_L\int_0^{t_b} \omega_m dt \tag{6.80}$$

Equation (6.80) shows that the energy loss is reduced, if braking takes place under a loaded condition.

6.13.3 Energy Loss during Speed Reversal

The counter current braking (plugging) is applied for speed reversal of a dc motor. The supply lines to the armature only are interchanged. The applied voltage becomes $-V$. If counter current braking is applied to the motor at no-load, we have

$$\omega_{in} = \omega_0, \text{ and } -\omega_{fin} = \omega_0$$

Equation (6.74) may be referred to. On the basis of this equation, we obtain

$$A_{ccb} = \int_{\omega_0}^{-\omega_0} J(-\omega_0 - \omega_m)d\omega_m = J\int_{-\omega_0}^{\omega_0} (\omega_0 + \omega_m)\, d\omega_m = 2J\omega_0^2 \tag{6.81}$$

The complete reversal operation can be divided into two parts. The first part consists of braking to standstill for which energy loss is $3/2J\omega_0^2$ and the second part consists of acceleration to no-load speed in the reverse direction for which energy loss is $1/2J\omega_0^2$. It may be noted that, while braking to standstill without load, the energy amounting to $1/2J\omega_0^2$ is obtained from the rotating parts of the motor and the balance amount is drawn from the supply.

On the other hand, while braking the motor to standstill with load torque acting, the energy loss during reversal is given by

$$A_{ccb} = \int_{\omega_L}^{0} -J(\omega_0 + \omega_m)d\omega_m - \int_0^{t_b} T_L(\omega_0 + \omega_m)dt$$

$$= J\left[\omega_0\omega_L + \frac{1}{2}\omega_L^2\right] - T_L\int_0^{t_b} (\omega_0 + \omega_m)dt \tag{6.82}$$

Example 6.7

A 220 V dc series motor driving a constant load torque runs at 200 rad/sec and draws a current of 20 A from the supply. The total resistance of the armature and the field is 1.0 Ω. The moment of inertia of the motor together with load is 5 kg · m^2. Calculate the total energy dissipated in the armature circuit, if the motor starts from rest and attains the steady speed of 200 rad/sec within a time period of 2.5 seconds.

Solution

The equations for electrical and mechanical systems based upon usual notations are

$$E = V - iR = K\phi\omega = k_T i\omega$$

$$T = K\phi i = k_T i^2 = J\frac{d\omega_m}{dt} + T_L$$

or

$$i^2 = \frac{J}{k_T}\frac{d\omega_m}{dt} + \frac{T_L}{k_T}$$

The total energy dissipated in the armature circuit is given by the following equation:

$$\int_{t_1}^{t_2} i^2 R dt = \frac{JR}{k_T}\int_{\omega_1}^{\omega_2} d\omega_m + \frac{T_L R}{k_T}\int_{t_1}^{t_2} dt = \frac{JR}{k_T}(\omega_2 - \omega_1) + \frac{T_L R}{k_T}(t_2 - t_1)$$

$V = 220$ V, $I = 20$ A, $R = 1\ \Omega$, $J = 5\ \text{kg}\cdot\text{m}^2$, $\omega = 200$ rad/s

$$E = k_T I\omega = V - IR = 220 - (20 \times 1) = 200 \text{ V}$$

$$k_T = \frac{E}{I\omega} = \frac{200}{20 \times 200} = 0.05$$

$$T_L = k_T I^2 = 0.05 \times (20)^2 = 20 \text{ N}\cdot\text{m}$$

$$\omega_2 = \omega, \quad \omega_1 = 0, \quad t_2 - t_1 = 2.5 \text{ sec.}$$

Therefore,

$$\text{Energy loss} = \frac{JR}{k_T}(\omega_2 - \omega_1) + \frac{T_L R}{k_T}(t_2 - t_1)$$

$$= \frac{5 \times 1 \times 200}{0.05} + \frac{20 \times 1 \times 2.5}{0.05} = 21000 \text{ J}$$

6.14 DYNAMIC (TRANSIENT) RESPONSE OF INDUCTION MOTORS

In the year 1938, H.C. Stanley derived a set of five nonlinear differential equations for three-phase induction motors by two-axis transformation modelled on the work of Park. These equations are very complicated because of mathematical complexity involved. Later, in the year 1978, a set of three nonlinear equations in place of five nonlinear equations was derived, facilitating the computation. The method of computation has also been suggested. The novelty of this method is that the equations are applicable to an induction motor in association with inverter supply of any waveform, viz. sinusoidal, PWM, square, stepped, continuous or discontinuous. It is beyond the scope of this book to give, in detail, the transient behaviour of the induction motor, as it is highly mathematical.

However, the equation of motion of an induction motor can be solved conveniently by the graphical method used to determine the starting and braking time. This method is discussed assuming that a cage motor is driving a fan load. The speed-torque characteristics of the motor and the load are assumed to be known. In the first step, these two characteristics are plotted in the second quadrant as shown in Fig. 6.12.

The dynamic torque ($T - T_L$) is plotted in the same quadrant. The dynamic equation of motion is given by

$$T - T_L = J\frac{d\omega_m}{dt} \tag{6.83}$$

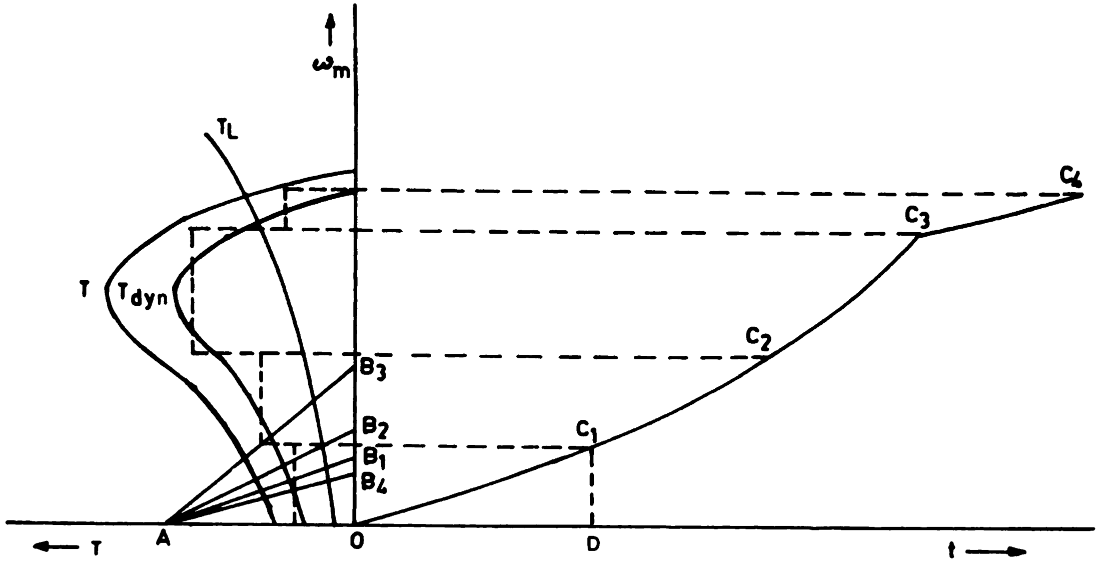


Fig. 6.12 Grapical method of determination of starting time of induction motors.

In the above equation, $d\omega_m$ and dt are replaced by finite increments of $\Delta\omega_m$, and Δt, respectively. Then

$$T - T_L = J\frac{\Delta\omega_m}{\Delta t} \tag{6.84}$$

The dynamic torque curve is replaced by the stepped torque curve; $T_{dyn}(T - T_L)$ is assumed to remain constant in each step; $\Delta\omega_m$ and Δt are speed and time increments after each step.

Torque and speed scales have already been chosen. For construction of the speed curve $\omega = f(t)$, the inertia and time scale factors should also be fixed up. The following relationship may be used:

$$\frac{\text{Torque scale factor}}{\text{Moment of inertia scale factor}} = \frac{\text{Speed scale factor}}{\text{Time scale factor}} \tag{6.85}$$

So, if now the moment of inertia scale is chosen, the time scale is fixed up on the basis of the above relationship and vice versa. The length OA is equal to the moment of inertia to scale. It may be noted that with an increase in the number of steps, the accuracy obtained is greater. For simplicity only, four steps have been taken (Fig. 6.12). The cut-offs OB_1, OB_2, OB_3 and OB_4, which have been marked out on the ordinate, are equal to the dynamic torque of the steps 1, 2, 3, and 4, respectively. The points B_1, B_2, B_3 and B_4 are joined to A by straight lines. The lines OC_1, C_1C_2, C_2C_3 and C_3C_4 are drawn parallel to AB_1, AB_2, AB_3, and AB_4, respectively. The curve $OC_1C_2C_3C_4$ represents $\omega_m = f(t)$. This can be explained in the following way:

$$\frac{OB_1}{OA} = \frac{C_1D}{OD}$$

where $OB_1 = T - T_L$, $OA = J$, $C_1D = \Delta\omega_1$.

Then, according to Eq. (6.84), OD is equal to Δt to scale.

The starting time can also be determined by grapho-analytical integration of the equation of motion. Equation (6.84) may be written as

$$\Delta t = \frac{J\Delta\omega_m}{T - T_L} \tag{6.86}$$

In this method too, the dynamic torque ($T - T_L$) is plotted in the second quadrant as in the previous method. The range of speed up to the stable operating condition is divided into a series of speed increments. The dynamic torque ($T - T_L$) in any step is the mean torque in that step. Then, by using Eq. (6.86), the time increment can be determined. The step-by-step integration can be done to find the run-up time (Fig. 6.13).

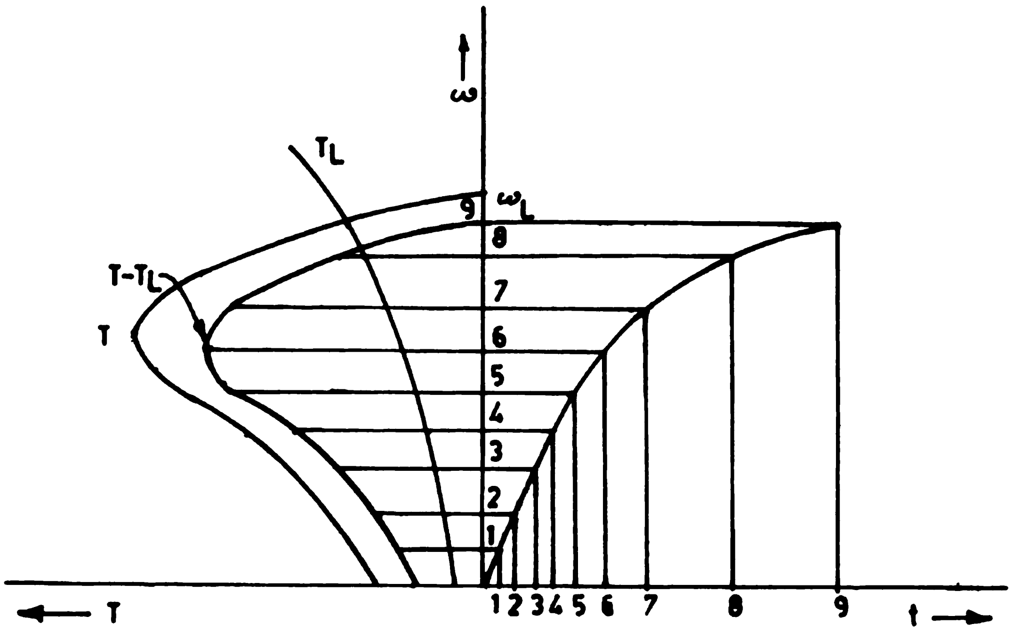


Fig. 6.13 Graphical integration method of determination of starting time of induction motors.

If the speed increments are taken to be equal for each step, then the run-up (starting) time up to any step m is given by

$$t = J\Delta\omega_m \sum_{1}^{m} \frac{1}{(T - T_L)} \tag{6.87}$$

Thus, $\omega = f(t)$ can be determined by the above methods and $T = f_1(t)$ can be determined by using $\omega = f(t)$. If $T = f_1(t)$ is known, then $i = f_2(t)$ can also be determined. The above methods can be used to determine braking time, dynamic and transient behaviour of induction motors. However, electrical transients in induction motors may be ignored in many cases, because of their short duration compared to mechanical transients.

Example 6.8

A three-phase, 440 V, 40 kW, 50 Hz induction motor driving a fan, whose torque is proportional to the square of the speed, has the speed-torque characteristic given below:

Speed (rpm)	2950	2750	2500	2400	2250	2000	1750	1500	1000	500	0
Torque (N · m)	120	288	360	369	360	297	252	207	162	144	150

Determine the run-up time of the fan motor having rated torque of 120 N · m to accelerate to its rated speed of 2950 rpm. The total inertia of the motor and the fan is equal to 2.0 kg · m^2.

Solution

From the data supplied, Table 6.1 is prepared to obtain the speed-time relationship.

$$J = 2 \text{ kg} \cdot \text{m}^2, \qquad T_L(\text{rated}) = 120 \text{ N} \cdot \text{m}, \qquad N(\text{rated}) = 2950 \text{ rpm}$$

$$\Delta t = \frac{J\Delta\omega_m}{T - T_L}, \qquad \text{Fan torque } T_L = \frac{T_L(\text{rated})}{N^2(\text{rated})} \cdot N^2$$

Speed change of a step is assumed.

$$t = \text{run-up time to full-load speed} = 3.475 \text{ sec.}$$

Table 6.1 Speed-time Relation

Speed (rpm) *at the start of a step*	*Motor torque* (N · m) T_e	*Fan torque* (N · m) T_L	$1/T_e - T_L$	$\Delta\omega_m$ (rad/s)	Δn (rps)	*Speed* (rpm) *at the end of a step*	Δt (sec)	t(sec)
0	150	0.00	0.0066	52.3	8.33	500	0.69	0.69
500	144	3.45	0.007	52.3	8.33	1000	0.73	1.42
1000	162	13.80	0.0067	52.3	8.33	1500	0.7	2.12
1500	207	31.00	0.0057	26.15	4.165	1750	0.3	2.42
1750	252	41.50	0.0047	26.15	4.165	2000	0.246	2.666
2000	297	55.20	0.0041	26.15	4.165	2250	0.214	2.88
2250	360	69.00	0.0034	15.7	2.5	2400	0.106	2.986
2400	369	79.50	0.00345	10.46	1.67	2500	0.072	3.06
2500	360	86.20	0.00365	26.15	4.165	2750	0.19	3.25
2750	288	104.30	0.00354	20.9	3.34	2950	0.225	3.475

Alternatively, the dynamics of starting of an induction motor, especially for high inertia loads, can be studied in the following way.

6.14.1 Starting an Induction Motor with Pure Inertia Load

Let us suppose that

(a) there is no friction and the motor torque is used for acceleration only, and

(b) the motor is started direct-on-line (DOL), i.e. in a single step.

The equation of motion may be written as

$$T_e = \frac{2T_m}{\left(\frac{s}{s_m} + \frac{s_m}{s}\right)} = J\frac{d\omega_m}{dt} \tag{6.88}$$

Now

$$\omega_m = \omega_s(1 - s)$$

or

$$\frac{d\omega_m}{dt} = -\omega_s\frac{ds}{dt}$$

Then

$$\frac{2T_m}{\left(\frac{s}{s_m} + \frac{s_m}{s}\right)} = -J\omega_s\frac{ds}{dt}$$

or

$$dt = -\frac{J\omega_s}{2T_m}\left(\frac{s}{s_m} + \frac{s_m}{s}\right)ds = -\frac{T_{em}}{2}\left(\frac{s}{s_m} + \frac{s_m}{s}\right)ds \tag{6.89}$$

where T_{em} = electromechanical time constant, which is defined as the time required by a drive with moment of inertia J to accelerate to the synchronous speed ω_s with maximum torque T_m. That is

$$T_{em} = \frac{J\omega_s}{T_m}$$

Acceleration from standstill ($s = 1$) to slip s gives

$$t_{st} = -\frac{T_{em}}{2}\int_1^s\left(\frac{s}{s_m} + \frac{s_m}{s}\right)ds = \frac{T_{em}}{2}\left[\frac{(1-s^2)}{2s_m} - s_m \log_e s\right]$$

The normal operating slip of an induction motor is about 0.05.

The starting time of an induction motor with pure inertia is thus given by

$$t_{st} = \frac{T_{em}}{2}\left[\frac{(1-(0.05)^2)}{2s_m} - s_m \log_e (0.05)\right]$$

or

$$\frac{t_{st}}{T_{em}} = \frac{1}{4s_m} + 1.5s_m \tag{6.90}$$

It shows that the starting time (per unit) depends on s_m. If this relationship is plotted for various values of s_m, it is found that for the quickest start, s_m should be between 0.4 to 0.5 (strictly 0.407); s_m = 0.407 is, however, uneconomical. Thus

$$\left(\frac{t_{st}}{T_{em}}\right)_{\min} = 1.22 \tag{6.91}$$

Rotor heating

$$I_2' = \frac{sE_1}{\sqrt{(r_2')^2 + (sx_2')^2}} = \frac{E_1}{x_2'\sqrt{1+\left(\frac{s_m}{s}\right)^2}} = \frac{(I_2')_{bl}}{\sqrt{1+\left(\frac{s_m}{s}\right)^2}}$$

$$I_2 = \frac{(I_2)_{bl}}{\sqrt{1+\left(\frac{s_m}{s}\right)^2}}$$

$$\text{Rotor heat loss per phase during starting} = \int_0^{t_{st}} I_2^2 r_2 dt = \int_1^s -\frac{(I_2)_{bl}^2 r_2 s^2}{s^2+s_m^2}\left(\frac{J\omega_s}{2T_m}\right)\left(\frac{s}{s_m}+\frac{s_m}{s}\right)ds$$

$$= \frac{J\omega_s}{4T_m}\cdot\frac{(I_2)_{bl}^2\cdot r_2}{s_m}(1-s^2) \tag{6.92}$$

During acceleration from rest ($s = 1$) to very small slip such that $s^2 = 0(s \cong 0)$, we have

$$\text{Rotor heat loss per phase} = \frac{(I_2)_{bl}^2\cdot r_2}{4s_m}\cdot\frac{J\omega_s}{T_m} \tag{6.93}$$

The maximum torque T_m in synchronous watts per phase (N · m) $= \dfrac{E_2(I_2)_{bl}}{2\omega_s}$

and

$$s_m = \frac{(I_2)_{bl}\cdot r_2}{E_2}$$

So, the rotor loss per phase during starting is

$$W_2 = \frac{(I_2)_{bl}^2\cdot r_2\cdot E_2}{4(I_2)_{bl}\cdot r_2}\cdot\frac{J\omega_s\cdot 2\omega_s}{E_2\cdot(I_2)_{bl}} = \frac{1}{2}J\omega_s^2 \tag{6.94a}$$

The energy loss in the rotor circuit in starting from rest is equal to the kinetic energy stored. This loss is unavoidable. However, the loss within the rotor circuit can be reduced by transferring a part of it to an external resistance. This is possible only for slip-ring induction motors. Heat loss equal to $W_2(r_1/r_2')$ occurs in the stator circuit. Inadequate torque may result in excessive run-up time. If a slip-ring induction motor is started with an external resistance in the rotor circuit, the rotor current at a given speed is reduced and thereby the accelerating torque is also reduced and a delayed start takes place. Hence, the energy loss remains constant in spite of the fact that the rotor current is decreased.

6.14.2 Induction Motor Started on Load

The dynamic equation of motion is

$$T_e - T_L = T_{dyn} = J\frac{d\omega_m}{dt}$$

or

$$dt = J\frac{d\omega_m}{(T_e - T_L)}$$

where T_e = developed torque, T_L = load torque. Also

$$\omega_m = \omega_s(1 - s)$$

or

$$d\omega_m = -\,\omega_s\,ds$$

Energy loss in the rotor circuit during starting $= 3\int_0^{t_{st}} I_2^2 r_2 dt = \omega_s \int_0^{t_{st}} sT_e dt = -J\omega_s^2 \int_{s_{in}}^{s_{fin}} \frac{T_e}{(T_e - T_L)} s\,ds$

If the motor is started from rest to rated speed, the energy loss is

$$J\omega_s^2 \int_{s_{fin}}^{1} \frac{T_e}{(T_e - T_L)} s\,ds \tag{6.94b}$$

Equation (6.94b) shows that the energy loss during starting of an induction motor on load, increases compared to energy loss during starting without load. At no-load, the energy loss does not depend upon the nature of the speed-torque characteristic curve. The T_e and T_L are functions of speed. The energy loss during starting of an induction motor is minimum with direct-on-line start. The T_e is maximum as full voltage is applied. With reduced voltage starting, the energy loss increases.

6.15 COUNTER CURRENT BRAKING OF INDUCTION MOTORS

In counter current braking (plugging), we have

$$s = \frac{\omega_s + \omega_m}{\omega_s}$$

It is assumed that friction is negligible and braking takes place without load. Thus, we have

$$\frac{2T_m}{\left(\dfrac{s}{s_m}+\dfrac{s_m}{s}\right)} = J\frac{d\omega_m}{dt} = J\omega_s\frac{ds}{dt}$$

or

$$dt = \frac{J\omega_s}{2T_m}\left(\frac{s}{s_m}+\frac{s_m}{s}\right)ds = \frac{T_{em}}{2}\left(\frac{s}{s_m}+\frac{s_m}{s}\right)ds \tag{6.95}$$

Also

$$t_{ccb} = \frac{T_{em}}{2}\int_{s_{in}}^{s_{fin}}\left(\frac{s}{s_m}+\frac{s_m}{s}\right)ds$$

Now

$$s_{in} = 2 - s \cong 2 \qquad \text{and} \qquad s_{fin} = 1$$

Therefore

$$t_{ccb} = \frac{T_{em}}{2}\int_{2}^{1}\left(\frac{s}{s_m}+\frac{s_m}{s}\right)ds$$

or

$$\frac{t_{ccb}}{T_{em}} = 0.345 s_m + \frac{0.75}{s_m} \tag{6.96}$$

For quickest braking, $s_m = 1.47$, for which $\dfrac{t_{ccb}}{T_{em}} = 1.027$.

6.15.1 Rotor Heating

Rotor loss per phase $= \displaystyle\int_0^{t_{ccb}} I_2^2 r_2\, dt$

Since

$$I_2 = \frac{(I_2)_{bl}}{\sqrt{1+\left(\dfrac{s_m}{s}\right)^2}}$$

Therefore

$$(W_2)_{ccb} = \int_{s_{in}}^{s_{fin}} \frac{(I_2)_{bl}^2 r_2 s^2}{s^2+s_m^2}\cdot\frac{J\omega_s}{2T_m}\left(\frac{s}{s_m}+\frac{s_m}{s}\right)ds$$

Now

$$s_{in} = 2 \quad \text{and} \quad s_{fin} = 1$$

Therefore

$$(W_2)_{ccb} = \frac{J\omega_s}{4T_m} \cdot \frac{(I_2)_{bl}^2 r_2}{s_m} [s^2]_2^1 = -\frac{3J\omega_s}{4T_m} \cdot \frac{(I_2)_{bl}^2 r_2}{s_m} \tag{6.97}$$

It may be noted that T_m is negative. Hence, the rotor loss is positive.

The maximum torque in synchronous watts per phase $T_m = -\frac{1}{2} E_2(I_2)_{bl}$.

Therefore

$$(W_2)_{ccb} = \frac{-3 \cdot (I_2)_{bl}^2 r_2 E_2}{4(I_2)_{bl} r_2} \cdot \frac{J\omega_s 2\omega_s}{-E_2(I_2)_{bl}} = \frac{3}{2} J\omega_s^2 \tag{6.98}$$

Stator loss during counter current braking $= (W_2)_{ccb} \frac{r_1}{r_2'}$.

Hence, the total energy loss during counter current braking up to standstill

$$= \frac{3}{2} J(\omega_s)^2 \left(1 + \frac{r_1}{r_2'}\right) \tag{6.99}$$

Energy loss (Joules) during reversal $= 2J\omega_s^2 \left(1 + \frac{r_1}{r_2'}\right)$ (6.100)

6.16 DYNAMIC BRAKING OF INDUCTION MOTORS

The equation of motion during dynamic braking is given by

$$-\frac{2T_m}{\left(\frac{S}{S_m} + \frac{S_m}{S}\right)} = J\frac{d\omega_m}{dt} \tag{6.101}$$

Dynamic braking slip $S = \frac{\omega_m}{\omega_s}$

or

$$\frac{d\omega_m}{dt} = \omega_s \frac{dS}{dt}$$

Then

$$-\frac{2T_m}{\left(\frac{S}{S_m} + \frac{S_m}{S}\right)} = J\omega_s \frac{dS}{dt}$$

or

$$dt = -\frac{J\omega_s}{2T_m}\left(\frac{S}{S_m}+\frac{S_m}{S}\right)dS = -\frac{T_{em}}{2}\left(\frac{S}{S_m}+\frac{S_m}{S}\right)dS$$

Also

$$t_{db} = -\frac{T_{em}}{2}\int_{s_{in}}^{s_{fin}}\left(\frac{S}{S_m}+\frac{S_m}{S}\right)dS$$

For $S_{in} = 1$ and $S_{fin} = 0.05$, we have

$$t_{db} = T_{em}\left(1.5S_m + \frac{1}{4S_m}\right)$$

or

$$\frac{t_{db}}{T_{em}} = \frac{1}{4S_m} + 1.5\,S_m \tag{6.102}$$

The expression for t_{db} in Eq. (6.102) is similar to that of t_{st} in Eq. (6.90).
Hence, the minimum time of braking is

$$\frac{t_{db}}{T_{em}} = 1.22, \text{ at } S_m = 0.407 \tag{6.103}$$

Example 6.9

A 3.73 kW, 7.8 A, three-phase, 50 Hz, 1450 rpm, 0.84 power factor, delta-connected cage induction motor is to undertake frequent starts. The moment of inertia of the drive is 0.5 kg · m^2. Assume the ratio of stator to rotor resistance to be unity. Neglect magnetizing current and rotational losses.

(a) Determine the number of starts per minute that this drive can make under no-load conditions.

(b) How many starts per minute can be made, if the rated speed were 960 rpm, all the other parameters remaining unaltered?

Solution

$P_o = 3.73$ kW, $V = 415$ V, $I = 7.8$ A, power factor ($\cos\phi$) = 0.84

$N_m = 1450$ rpm, $f = 50$ Hz, $r_1 = r_2'$

$$p = \frac{120f}{N_s} \approx \frac{120f}{N_m} = \frac{120\times 50}{1450} = 4.138 \cong 4 \text{ (even no.)}$$

$$\omega_s = \frac{2\pi}{60}N_s = \frac{2\pi}{60}\cdot\frac{120f}{p} = \frac{2\pi\times 120\times 50}{60\times 4} = \frac{2\pi\times 1500}{60} = 157.1 \text{ rad/s}$$

Input power $P_i = \sqrt{3}VI \cos \phi = \sqrt{3} \times 415 \times 7.8 \times 0.84 = 4710$ W

Losses at full load $P_i - P_o = 4710 - 3730 = 980$ W

Energy loss per minute under full-load = 980 × 60 = 58800 Joules

(a) Energy loss during starting under no-load $= \frac{1}{2}J\omega_s^2\left(1+\frac{r_1}{r_2'}\right)$

$$= 0.5 \times 0.5 \times (157.1)^2 \times (1 + 1)$$

$$= 12340 \text{ Joules}$$

Therefore, the number of starts that can be made per minute $= \frac{58800}{12340} \approx 5$ starts

(b) If the motor speed (rated) $N_m = 960$ rpm, synchronous speed ω_s, with $p = 6$, as found by the procedure given above,

$$\omega_s = \frac{2\pi \times 120 \times 50}{60 \times 6} = \frac{2\pi \times 1000}{60} = 104.72 \text{ rad/s}$$

Energy loss during starting on no-load $= 0.5 \times 0.5 \times (104.72)^2 \times 2 = 5483$ Joules

The number of starts that can be made per minute $= \frac{58800}{5483} \approx 11$ starts

Example 6.10

A three-phase, 415 V, six-pole, 50 Hz, star-connected slip-ring induction motor has a sum of stator and rotor leakage reactance referred to the stator of 1.0 Ω. It is connected to a balanced 415 V supply and drives a pure inertia load. The moment of inertia of the rotor including the load is 10 kg · m². Direct-on-line starting is used and the rotor circuit resistance is adjusted, so that the motor brings its load from rest to 0.95 of the synchronous speed in the shortest possible time. Calculate the value of the rotor resistance referred to the stator and the minimum time to reach 0.95 of synchronous speed.

Solution

$f = 50$ Hz, $\quad p = 6, \quad J = 10 \text{ kg} \cdot \text{m}^2, \quad (x_1 + x_2') = 1\ \Omega$

$$\omega_s = \frac{2\pi}{60}\cdot\frac{120f}{p} = \frac{2\pi \times 120 \times 50}{60 \times 6} = \frac{2\pi \times 1000}{60} = 104.72 \text{ rad/s}$$

r_1 is assumed to be small.

$$V_{ph} = \frac{415}{\sqrt{3}} = 240 \text{ V}$$

$$T_m = \frac{3V_{ph}^2}{2\omega_s(x_1 + x_2')} = \frac{3 \times (240)^2}{2 \times 104.72 \times 1.0} = 825 \text{ N} \cdot \text{m}$$

$$T_{em} = \frac{J\omega_s}{T_m} = \frac{10 \times 104.72}{825} = 1.27 \text{ sec.}$$

Using Eq. (6.91), $\left(\frac{t_{st}}{T_{em}}\right)_{\min} = 1.22$, we have

$$(t_{st})_{\min} = 1.22 \times 1.27 = 1.55 \text{ sec.}$$

Now

$$\frac{r_2'}{(x_1 + x_2')} = 0.407$$

Therefore

$$r_2' = 0.407\ \Omega$$

6.17 REDUCTION OF ENERGY LOSSES DURING STARTING

In the previous sections, it has been found that loss of energy during starting of an induction motor depends on the moment of inertia. The energy losses increase with increase in moment of inertia. When frequent starting is required as, for example, in rolling mill drives, smaller moment of inertia of the driving motor is called for in comparison to that of a general purpose motor. Smaller inertia can be achieved by increasing the length and decreasing the diameter of the motor such that power rating and speed of rotation are maintained constant. Alternatively, the inertia can be reduced by replacing a single motor with two motors of one-half the power rating. In this way, though inertia is reduced, the total weight is increased.

The energy loss in starting resistance of dc motors can be eliminated by employing an adjustable voltage system, thus dispensing with the external starting resistance. That the adjustable voltage method is energy-efficient, is illustrated in Fig. 6.14.

The supply voltage V is constant. If the armature current I_a is assumed to be constant during the starting period, then the

(a) input power remains constant as represented by AB (Fig. 6.14a).

(b) speed of rotation, back emf and useful power increase linearly with time as represented by OC.

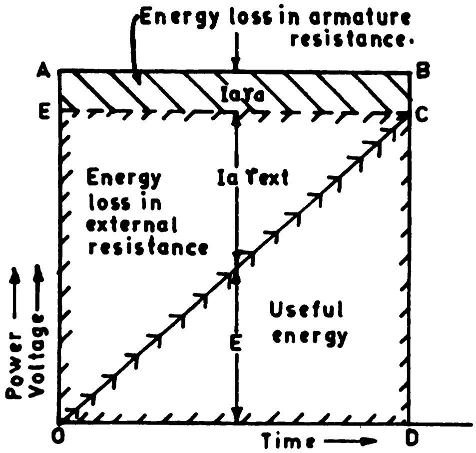
In rheostatic starting, the various related energies may be accounted for as under:

(a) Energy input = Area $ABDO$

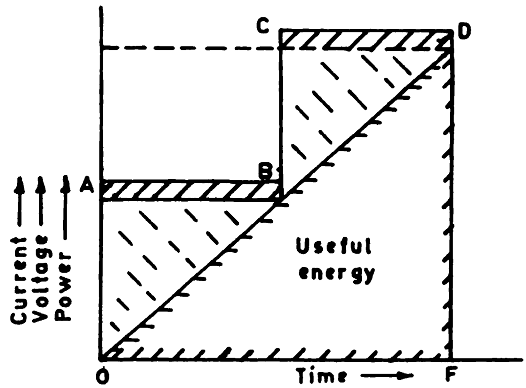
(b) Energy output = Area $OCDO$

(c) Energy loss in the external resistance = Area $OECO$

(d) Energy loss in the armature resistance = Area $ABCE$



(a) Rheostatic starting.



(b) Series-parallel starting.

Fig. 6.14 Energy losses in dc motors.

It is clear from Fig. 6.14a that in rheostatic starting of dc shunt motors, the initial energy loss is very high. This drawback of rheostatic starting can be eliminated to a large extent by step-by-step variation of supply voltage as in an adjustable voltage control or series-parallel control. Energy saving in series-parallel control is illustrated in Fig. 6.14b.

The supply voltage and field excitation are assumed to be constant. Initially the armatures are connected in series and then in parallel. The voltage across each armature is equal to $V/2$ in series connection and V in parallel connection. The armature current remains constant to develop constant torque and thereby constant acceleration. The total current from the supply is equal to the armature current in series connection and double the armature current in parallel connection. The supply voltage remaining constant, the input power in parallel connection is double of that in series connection. In Fig. 6.14b, *AB* and *CD* represent the measure of input power and current in series, and parallel connections, respectively. The back emf increases along *OFH*. The useful energy is given by the area *OHK*. The energy loss in series-parallel starting is reduced to half of that incurred in rheostatic starting.

Let us suppose that supply voltage is increased in m steps during starting. Since the speed of a dc shunt motor is approximately proportional to the applied voltage, the speed increment in each step will be equal, if the supply voltage is varied in equal steps. Therefore, the loss of energy in m no. of steps is,

$$W_2 = m\left[\frac{1}{2}J\left(\frac{\omega_m}{m}\right)^2\right] = \frac{\frac{1}{2}J\omega_m^2}{m} \tag{6.104}$$

The energy loss decreases as the number of steps increases.

A case study is cited as an example for economy gained in reduced voltage starting of cage induction motors. An experiment was conducted to analyse the cases of frequent breakdown of high frequency motor-alternator sets comprising a 107 kW, 230 V, three-phase, cage induction motor and a 75 kW, single-phase, 8 kHz, 750 V inductor alternator in the heat treatment and forging shop of

MAMC, Durgapur. The motor was meant for use with a star/delta starter. In the absence of 230 V, three-phase supply, the motor was working on 415 V, three-phase supply with star-connection of the stator winding through the DOL starter as prescribed by the manufacturer. After a span of 2–3 years of service, the stator winding and sometimes the rotor bars of the cage motor were found to burn out frequently. A detailed investigation revealed that during starting, the motor drew current nearly ten times the steady state value over 10–12 seconds for building up synchronous speed of 3000 rpm. The failure was never found to occur during running operation, which continued over 7–8 hours at a stretch. From the nature of the recurring type of failure, it became evident that the reduced voltage starting of the motor was essential to protect the motor from burning out. The standard star/delta starter was not being used for reduced voltage starting, since 230 V, three-phase supply did not exist at MAMC, Durgapur. Reduced voltage starting of the induction motor through a three-phase autotransformer, was considered to be the only possible solution, which is described in Section 7.4.3 along with its circuit diagram (Fig. 7.7).

Though the method is known to be efficient and flexible with tap-changing facility, it has reduced power factor and the acceleration is not very smooth. A three-phase auto-transformer was designed with output voltage tappings at 80%, 60% and 40% of the line voltage to supply the reduced voltage to the cage induction motor over a predetermined interval settable by means of a timer. The disadvantage of transient stress generation due to open transition from auto-transformer to line voltage is avoided by using Korndorfer connection.

One of the possible reasons for which the motor was drawing an abnormally high value of starting current could have been the poor regulation of the supply with an unusual voltage dip at the motor terminals (the allowable voltage dip is 5%). An oscillographic study of the transient behaviour of the motor terminal voltage under the DOL starting condition clearly showed that the terminal voltage of the motor before/during/after starting remained practically unaltered suggesting that there was no voltage dip in the supply system.

With DOL starting, the motor was drawing the maximum current which in about 12 sec. dropped to its normal no-load value of 60 A, on reaching the steady state speed. From the oscillogram, the DOL starting current was computed to be 1560 A, when the full-load current was specified to be 180 A. In auto-transformer starting (ATS), the motor was switched on through the 80% tapping point. After the set timing of 12 sec., the motor was switched on to the DOL connection. The oscillogram was taken again. The starting current was computed to be 1010 A. At the point of transition from ATS to DOL the motor current had shot up to 1150 A, which gradually dropped down to no-load value in less than 5 seconds. The total time for building up to rated speed was enhanced from 12 sec. in DOL starting to 17 sec. in auto-transformer starting (ATS). After plotting the heating curves, it was found that the value of thermal stress generation was substantially (35%) reduced in auto-transformer starting (ATS) compared to DOL starting.

The energy loss is reduced, when an induction motor is started by smooth variation of supply frequency under constant Volt/Hz operation. Both induction and dc shunt motors have exactly identical torque-speed characteristics. The Ward-Leonard method of dc shunt motors and constant Volt/Hz operation of induction motors are of the same type. If the supply voltage is increased in m equal steps maintaining constant Volt/Hz ratio, such that the step changes in speed are equal, the energy loss during starting is given by Eq. (6.104).

Two-speed starting of pole-changing induction motors is more efficient than single-step starting. A pole-changing induction motor having 2 : 1 speed ratio can be started on high voltage to maximum speed. Alternatively, it can be started initially on low-speed winding to half the speed and then to full speed on high-speed winding. In the alternative method of starting, the energy loss is halved.

The lower and upper synchronous angular velocities of the pole-changing induction motor are ω_{s1} and ω_{s2} respectively.

$$\omega_{s2} = 2\omega_{s1}$$

J = moment of inertia of the motor

(1) Total energy loss (Joules) in single-step starting to ω_{s2}

$$W_2(\text{direct}) = \frac{1}{2}J\omega_{s2}^2\left(1+\frac{r_1}{r_2'}\right) = 2J\omega_{s1}^2\left(1+\frac{r_1}{r_2'}\right) \tag{6.105}$$

(2) Energy loss in two-step starting

First step: Acceleration from rest to ω_{s1}

$$W_2(\textit{Step } 1) = \frac{1}{2}J\omega_{s1}^2\left(1+\frac{r_1}{r_2'}\right) \tag{6.106}$$

Second step: Acceleration from ω_{s1} to ω_{s2}

$$W_2(\textit{Step } 2) = \frac{1}{2}J(\omega_{s2}-\omega_{s1})^2\left(1+\frac{r_1}{r_2'}\right) = \frac{1}{2}J\omega_{s1}^2\left(1+\frac{r_1}{r_2'}\right) \tag{6.107}$$

The combined energy loss in two-step starting

$$W_2 = J\omega_{s1}^2\left(1+\frac{r_1}{r_2'}\right) \tag{6.108}$$

Therefore, two-step starting provides 50% energy saving.

(3) The energy loss in stepped braking

(a) Regenerative braking from ω_{s2} to ω_{s1}

$$\text{Energy loss} = \frac{1}{2}J(\omega_{s2}-\omega_{s1})^2\left(1+\frac{r_1}{r_2'}\right) = \frac{1}{2}J\omega_{s1}^2\left(1+\frac{r_1}{r_2'}\right) \tag{6.109}$$

(b) Counter current braking (plugging) from ω_{s1} to standstill

$$\text{Energy loss} = \frac{3}{4}J\omega_{s1}^2\left(1+\frac{r_1}{r_2'}\right) \tag{6.110}$$

(4) Energy loss in dynamic braking will always be equal to the stored energy at the time of its application.

(5) Energy loss in the stator of an induction motor is reduced by increasing the rotor resistance.

Example 6.11

An eight-pole, 50 Hz induction motor coupled to a flywheel drives a load, which requires a torque T_0 = 110 N · m, when running light. For an intermittent period of 8 seconds, a pulse load rising instantaneously to 550 N · m is supplied (Fig. 6.15). What must be the combined inertia of the system to ensure that the peak motor torque does not exceed 400 N · m? The motor characteristic may be taken as linear and giving a torque of 350 N · m at a slip of 5%. If the coupled inertia is changed to 200 kg · m^2, what will be the peak motor torque with the same duty?

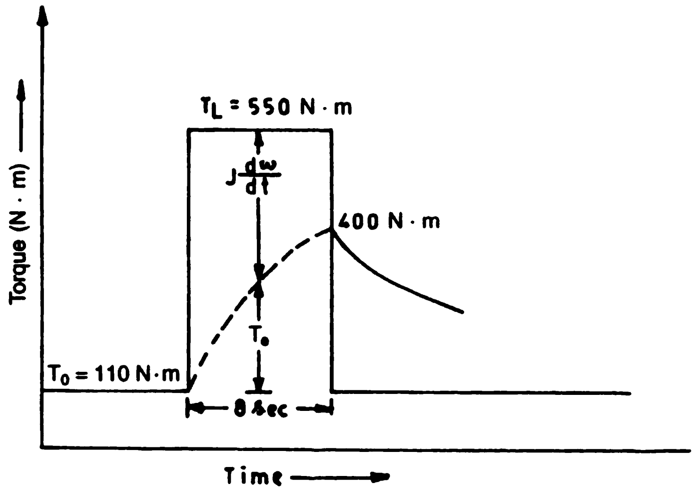


Fig. 6.15 Torque diagram of the induction motor of Example 6.11.

Solution

$f = 50$ Hz, $\quad p = 8, \quad T'_e = 350$ N · m, $\quad s' = 0.05$

$$\omega_s = \frac{2\pi}{60} \cdot \frac{120f}{p} = \frac{2\pi \times 120 \times 50}{60 \times 8} = \frac{2\pi \times 750}{60} = 78.54 \text{ rad/s}$$

Torque increases with increase of slip. Speed-torque characteristic is linear.

$$T_e = Ks$$

From the given data

$$K = \frac{T'_e}{s'} = \frac{350.0}{0.05} = 7000$$

The motor speed $\omega_m = (1 - s)\omega_s$

$$\frac{K}{\omega_s} = \frac{7000}{78.54} = 89.1$$

$$T_{em} = \frac{J\omega_s}{K} = \frac{J}{89.1}$$

The equation of motion is

$$T_e = T_L + J\frac{d\omega_m}{dt}$$

or

$$Ks = T_L - J\omega_s\frac{ds}{dt}$$

or

$$\frac{J\omega_s}{K}\frac{ds}{dt} + s = \frac{T_L}{K} \tag{6.111}$$

The solution is

$$s = s_0 + \left(\frac{T_L}{K} - s_0\right)\left(1 - e^{-\frac{K}{J\omega_s}t}\right) \tag{6.112}$$

or

$$s = \frac{T_L}{k} - \left(\frac{T_L}{K} - s_0\right)e^{-\frac{t}{T_{em}}}$$

Motor torque

$$T_e = T_0 + (T_L - T_0)\left(1 - e^{-\frac{K}{J\omega_s}t}\right)$$

or

$$T_e = T_L - (T_L - T_0)\,e^{-\frac{t}{T_{em}}}$$

Now

$$T_e = 400 \text{ N}\cdot\text{m}, \qquad T_L = 550 \text{ N}\cdot\text{m}, \qquad T_0 = 110 \text{ N}\cdot\text{m}, \qquad t = 8 \text{ sec.}$$

Therefore

$$400 = 110 + (550 - 110)\left(1 - e^{-\frac{8}{T_{em}}}\right)$$

or

$$400 = 550 - 440e^{-\frac{8}{T_{em}}}$$

Therefore

$$T_{em} = \frac{t}{\log_e\left(\dfrac{T_L - T_0}{T_L - T_e}\right)} = \frac{8}{\log_e\left(\dfrac{550-110}{550-400}\right)} = \frac{8}{1.076} = 7.434 \text{ sec.}$$

and

$$J = 89.1T_{em} = 89.1 \times 7.434 = 662.37 \text{ kg}\cdot\text{m}^2$$

With $J = 200\ \text{kg} \cdot \text{m}^2$

$$T_e = 110 + (550 - 110)\left(1 - e^{-\frac{8 \times 89.1}{200}}\right) = 538\ \text{N} \cdot \text{m}$$

PROBLEMS

6.1 A retardation test on a dc shunt motor gave the following results:

(a) With field not excited, the speed fell from 1530 rpm to 1470 rpm in 43 sec.

(b) With field normally excited, the same speed drop occurred in 26 sec.

(c) With an average load of 1.2 kW supplied by the armature, the same speed drop occurred in 20 sec.

Determine the moment of inertia of the rotating parts at 1500 rpm and the core loss for normal excitation at this speed.

6.2 The parameters of a separately excited dc motor are given below:

$$r_a = 1\ \Omega, \qquad L_a = 0, \qquad C = 1.8\ \text{V} \cdot \text{s/rad},$$

J (for motor and load combined) = $3.5\ \text{kg} \cdot \text{m}^2$,

F (for motor and load combined) = $0.3\ \text{kg} \cdot \text{m}^2/\text{sec}$

The load torque of the motor is proportional to the speed. The field current remains constant at rated value. A voltage of 220 V is suddenly applied across the motor armature terminals.

(a) Derive an equation for the motor speed as a function of time.

(b) Determine the steady-state speed.

(c) Determine the time required for the motor to reach 75% of the steady-state speed.

6.3 The following parameters are given for a separately excited dc motor:

240 V, 0.75 kW, 500 rpm, $r_a = 7.6\ \Omega$, $L_a = 0.06$ H,

$C = 4.23\ \text{V} \cdot \text{s/rad}$, $J = 0.14\ \text{kg} \cdot \text{m}^2$

Assume that load inertia equals the motor inertia. Find the transfer function indicating the natural frequency of oscillation and damping ratio. Also, find the speed response to a constant torque corresponding to a step change of 50 V in input voltage, using exact and approximate transformation.

6.4 A dc shunt motor rated at 200 V and 45 A has an armature resistance of 0.2 Ω. The shunt winding resistance is 200 Ω. The open circuit voltage of the machine running at 750 rpm with full excitation is 210 V. The motor is started from 200 V supply from which the field winding is already fully excited. An external resistor has been connected in the armature circuit to limit the maximum starting current to 91 A, and is left in the circuit, when the motor runs against the full-load torque. Find the equation for electromagnetic torque T_e as a function of speed. Develop the differential equation for the speed ω_m, when the total inertia is $11\ \text{kg} \cdot \text{m}^2$. Find the electromechanical time constant and the final steady-state speed.

6.5 A flywheel is fitted to the motor for which the specifications are given in Problem 6.4. The field current, the external resistance, and the mechanical torque T_L remain constant. The full voltage is applied. The speed reaches 250 rpm in 12 seconds. Determine the moment of inertia of the flywheel.

6.6 A dc motor with load coupled to the shaft is rotating at 275 rpm. The moment of inertia of the motor armature and the load is 900 kg · m^2, and the damping friction is 2000 N · m · s/rad. If the motor supply is cut off, find the time taken for the speed to drop to 100 rpm.

6.7 The two synchronous speeds of a pole-changing induction motor are 1500 and 3000 rpm. The rotor moment of inertia is 0.25 kg · m^2, and the stator to rotor resistance ratio is approximately equal to 1.5. Find

(a) the energy losses for direct starting of the motor in one and two steps without load ($T_L = 0$).

(b) the energy loss in one step from 3000 rpm to standstill by counter current braking, and

(c) the energy loss for stepped braking from 3000 to 1500 rpm by regenerative braking and from 1500 rpm to standstill by counter current braking.

6.8 Determine the time required to bring to rest by counter current braking, a load of moment of inertia 10 kg m^2 from a speed of 950 rpm by means of 440 V, six-pole, 50 Hz, star-connected, three-phase induction motor having the following parameters:

$$r_1 = 0, \quad r'_2 = 0.2\ \Omega, \quad x_1 = x'_2 = 0.5\ \Omega$$

What resistance per phase must be added to the rotor circuit to bring the motor to rest in the quickest time?

6.9 A separately excited dc motor is used to drive an inertia load of 1.5 kg · m^2. The motor inertia is 0.1 kg · m^2. The rotational loss, the armature inductance L_a and the frictional coefficient F are negligible. The armature resistance r_a is 0.2 Ω. The motor and the load run at a steady speed of 1200 rpm with rated field excitation and an armature voltage of 100 V. The armature terminal voltage is suddenly increased to 125 V. Determine the final steady-state speed of the motor-load system, after the step increase in the terminal voltage.

6.10 A separately excited dc motor has the following parameters:

$$C = 2.0\ \text{V} \cdot \text{s/rad}, \qquad r_a = 0.4\ \Omega, \qquad J = 4.5\ \text{kg} \cdot \text{m}^2$$

The rotational losses are negligible. The motor operates at no-load with armature terminal voltage of 220 V. The motor is braked to zero speed by counter current braking. Determine the time of braking.

CHAPTER 7

MOTOR STARTERS AND CONTROLLERS

7.1 DC MOTOR STARTERS

The schemes described in the following sections are used for starting dc shunt motors.

7.1.1 Starters Using Voltage Sensing Relays

The back emf or speed serves as a signal for operation of contactors which are used to decrease or switch off the starting resistance step-by-step. The armature or back emf provides the speed signal. Figure 7.1 shows the power and control circuits describing the principle of starting dc shunt motors as function of speed.

When the start push button is pressed, the relay M is energized closing the contactor M. The motor starts with resistances, R_1, R_2 and R_3 in the armature circuit. The back emf increases and armature current decreases. As the lower limit of starting current I_2 is reached, the relay 1AR picks up at a voltage, $V_{1A} = C\omega_1 + I_2 (R_a + R_1 + R_2 + R_3)$. The starting coil 1A is energized closing its contactor 1 A. The resistance R_1 is short circuited. Instantaneously, the armature current rises to the upper limit of the starting current and then decreases exponentially to its lower limit I_2, when the relay 2AR picks up at a voltage, $V_{2A} = C\omega_2 + I_2 (R_a + R_2 + R_3)$. The resistance R_2 is then short circuited. Similarly, R_3 is short circuited, when the relay 3AR picks up at a voltage, $V_{3A} = C\omega_3 + I_2 (R_a + R_3)$. It may be noted that $\omega_3 > \omega_2 > \omega_1$. V_{1A}, V_{2A} and V_{3A} are approximately equal. Though the accelerating relays have the same pick up voltages, they are actually speed sensitive.

7.1.2 Starters Using Current Sensing Relays

During starting of a dc motor, the current varies between two limits depending upon the number of steps, the load torque and the rate of acceleration required. Based on this principle, the starting resistance is short circuited step-by-step. As the armature current drops down to its lower value at any step, the relay acts to short circuit the particular sectional resistance (Fig. 7.2). At the beginning of a step, the armature current rises to the upper peak value and then exponentially decreases. At the upper peak value of current, the concerned accelerating relay contact opens. At the lower limit of current, the accelerating relay releases and the corresponding section of the starting resistance is short circuited. IR is the interlocking relay. Its operating time is greater than that of the

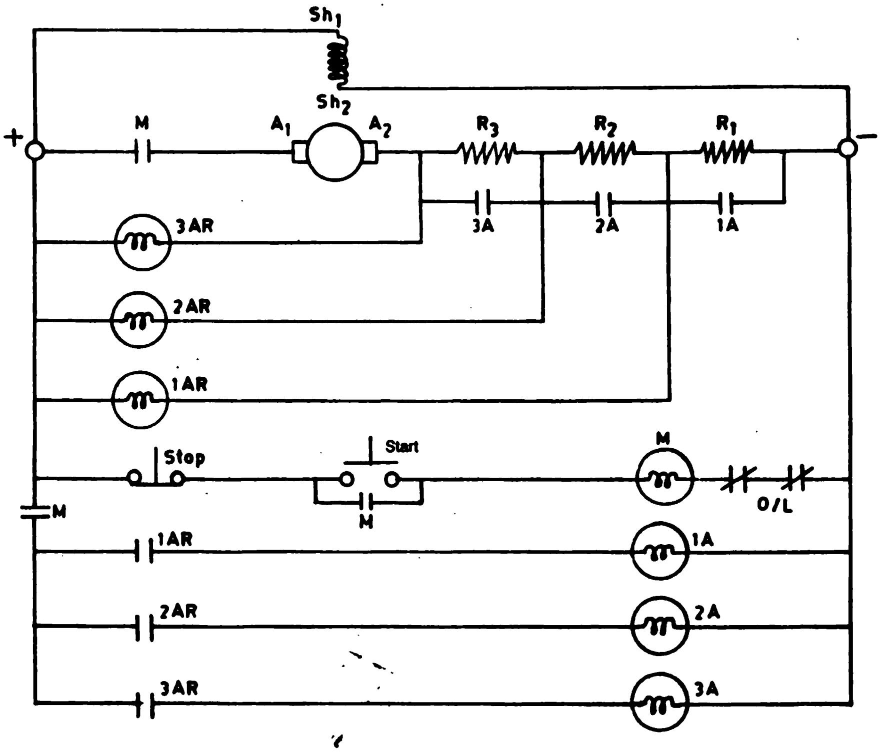


Fig. 7.1 DC shunt motor starter using voltage sensing relays.

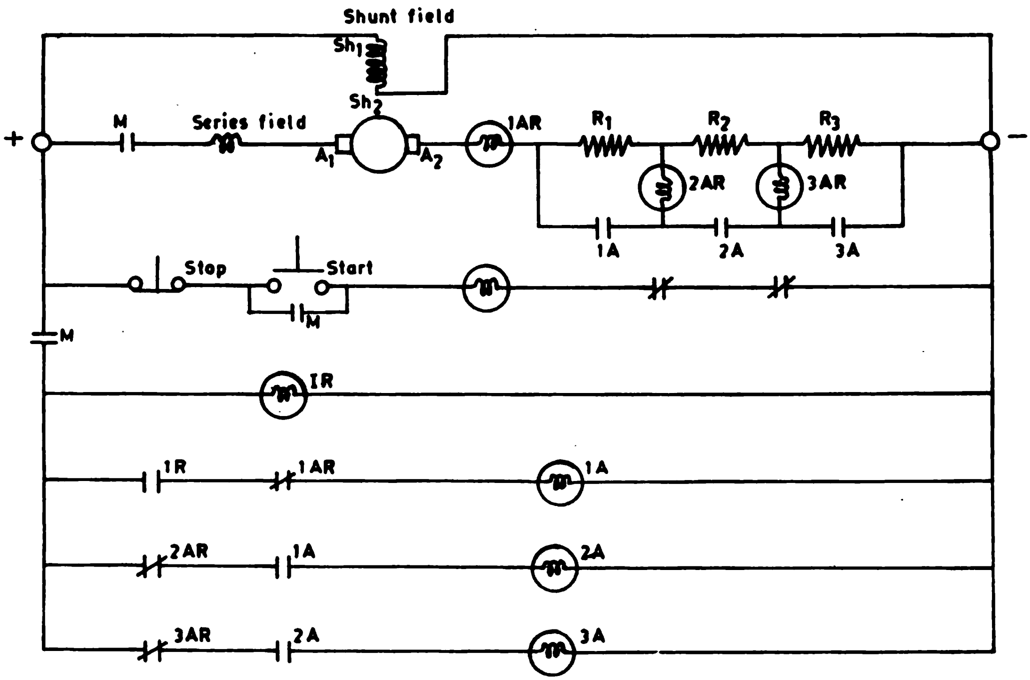


Fig. 7.2 DC shunt motor starter using current sénsing relays.

accelerating relay. The contacts, 1A, 2A and 3A close in turn, short circuiting the resistances, R_1, R_2, and R_3, in sequence.

7.1.3 Starters Using Time Delay Relays

The starting resistance can also be cut off at specific intervals of time by using time delay relays (timers). Normally, off delay timers are used. These timers pick up almost instantaneously on energization, but release with time delay, which is preset.

1AR, 2AR and 3AR are off delay timers, which are preset with required time delays. Before the start push button is pressed, the timer 1AR is already in energized state and its contacts, 1A, 2A and 3A are open (Fig. 7.3).

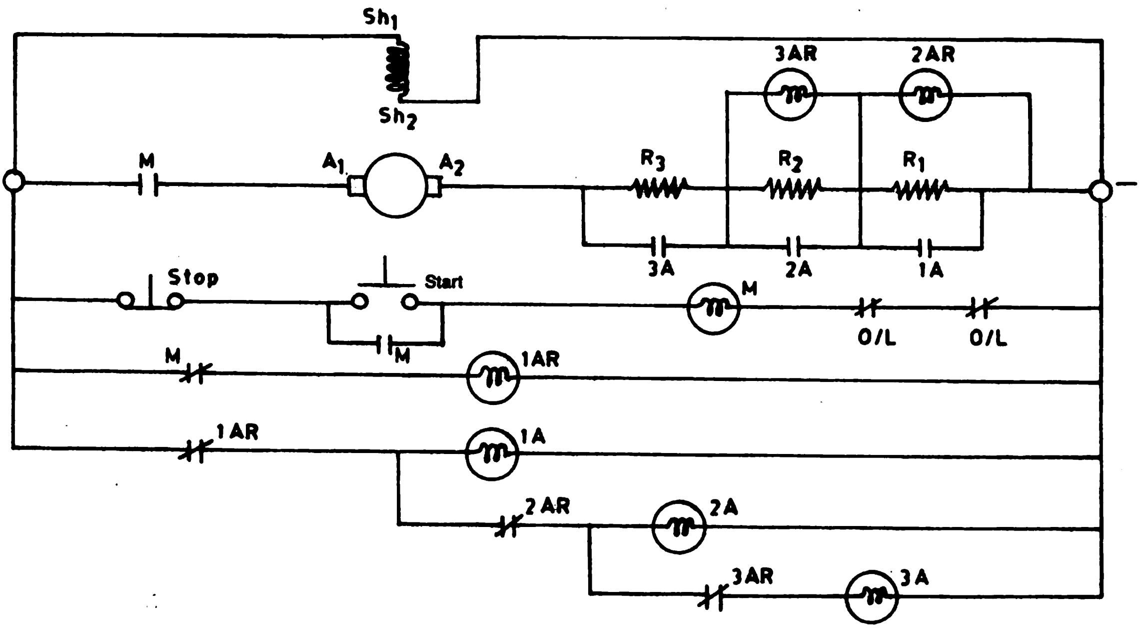


Fig. 7.3 DC shunt motor starter using time delay relays.

As soon as the start push button is pressed, the coil M is energized and the timer 1AR is de-energized. The motor starts with full resistance in the armature circuit. The contact 1AR drops out with elapse of its delay time. The coil 1A is energized closing its normally open contact 1A and simultaneously, the timer 2AR is de-energized. The contact 2AR drops out after its preset time delay, when the coil 2A is energized closing its contact 2A. The timer 3AR is now de-energized and similarly the contact 3A is closed. Thus, the starting resistance is short circuited step-by-step irrespective of the speed and current of the armature.

Electronic timers are used for automatic control of the starting cycle. A wide range of time delay settings from a few milliseconds to a few seconds can be obtained. Prolonged operation of the motor at a speed below normal is avoided in this method.

7.2 WOUND ROTOR INDUCTION MOTOR STARTERS BY VARIATION OF ROTOR RESISTANCE (MINE HOIST)

Wound rotor (slip ring) induction motors with rheostatic speed control are used, when these motors drive intermittent duty loads such as crane, wagon tripler, skip hoist, mine hoist, lift, etc.

7.2.1 Starters Using Time Delay Relays

Initially the off delay timers, 1TR, 2TR and 3TR are all energized and the coils, 1A, 2A, 3A and M are all de-energized (Fig. 7.4). As soon as the start push button is pressed, the coil M is energized.

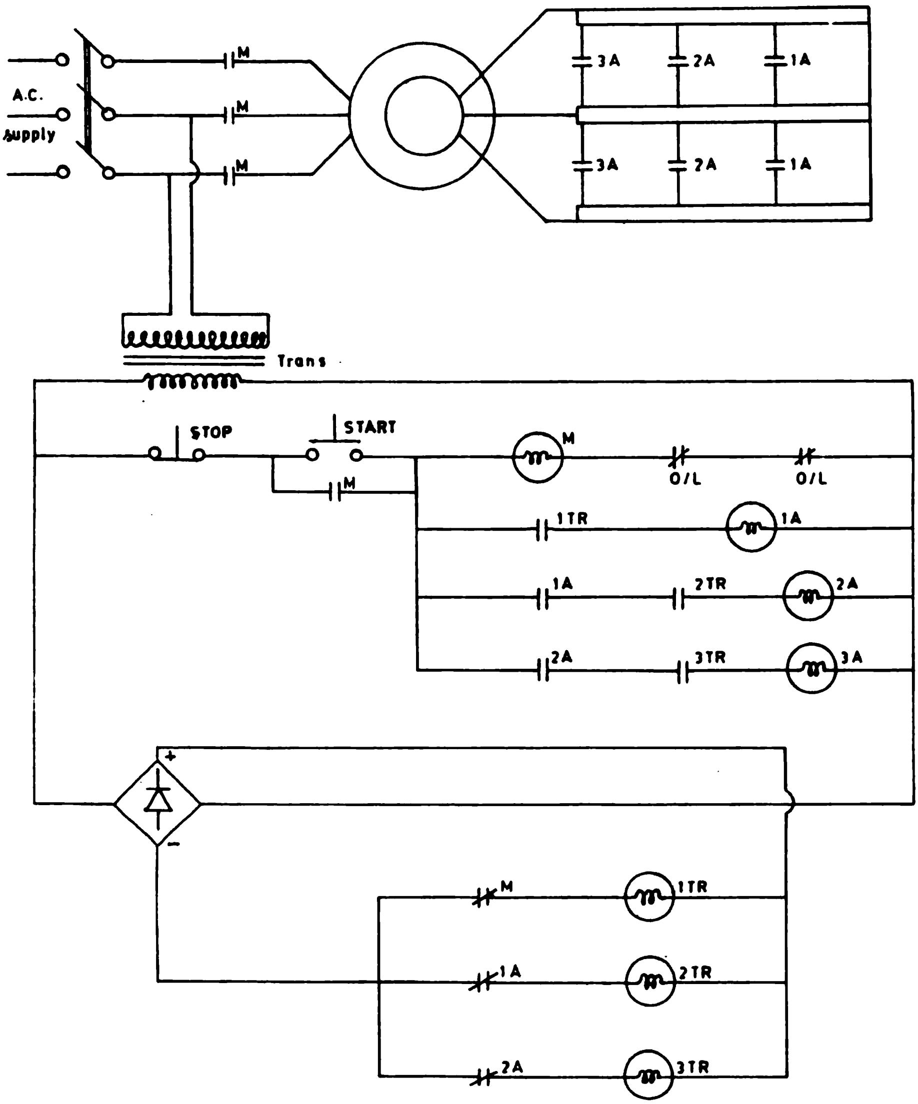


Fig. 7.4 Wound rotor induction motor starter using time delay relays.

The line contacts M are closed, thus supplying the voltages to the stator windings. The motor accelerates with full resistance in the rotor circuit. The blocking of starting push button is effected by a contact (NO) of M, which also switches the supply to the relay circuits at the same time. The contact (NC) of M in the timer circuit opens. The timer, 1TR drops out after its preset time delay. The coil 1A is energized, thus closing 1A contacts in the rotor circuit. On the other side, 1A contact in the timer circuit is opened at the same time. When the timer 2TR is de-energized after preset time delay, the coil 2A is energized, thus closing the contact 2A in the rotor circuit. Thereafter, the contacts 3A are closed in the rotor circuit after elapse of the delay time of 3TR. In this way, the starting resistance is short circuited in the rotor circuit in sequence.

7.2.2 Starters Using Frequency Sensing Relays

The other types of relays using rotor frequency or current sensing can also be used instead of time delay relays. The scheme using frequency sensing relays is described below (Fig. 7.5).

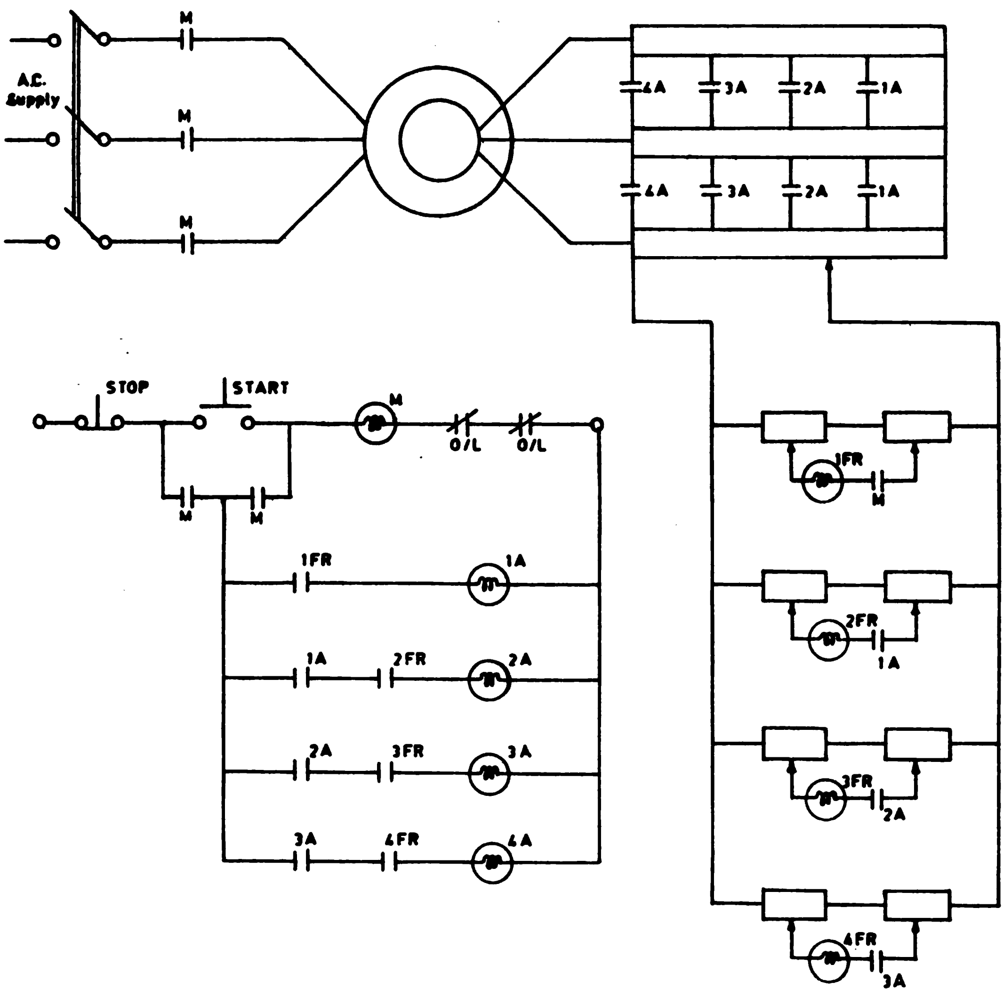


Fig. 7.5 Wound rotor induction motor starter using frequency sensing relays.

When the start push button is pressed, the line contactor M is energized closing its M contacts (NO) and starting the motor. Two M contacts (NO) block the start push button after its release, and this also energizes the main relay circuits.

As the rotor speeds up, its voltage and frequency drop. At a preset value corresponding to a particular rotor speed, the first frequency relay 1FR drops out, when its contact 1FR closes. This causes the contactor 1A to pick up and thus, the first set of rotor resistance is short circuited. Then the second frequency relay drops out closing its contact 2FR. This causes the relay 2A to pick up. Consequently, its contacts 2A in the rotor circuit short circuit the second set of rotor resistance. This process continues, until the last set of rotor resistance is short circuited. At any instant, when the motor is running, if the stop push-button is pressed, the coil M is de-energized, thus switching off the motor supply.

7.3 INDUCTION MOTOR STARTERS

7.3.1 DOL Starter with Provision for Speed Reversal

The direct-on-line (DOL) starter (Fig. 7.6) with provision for speed reversal is described. The forward push button FOR is pressed. The forward contactors F pick up providing proper supply to the

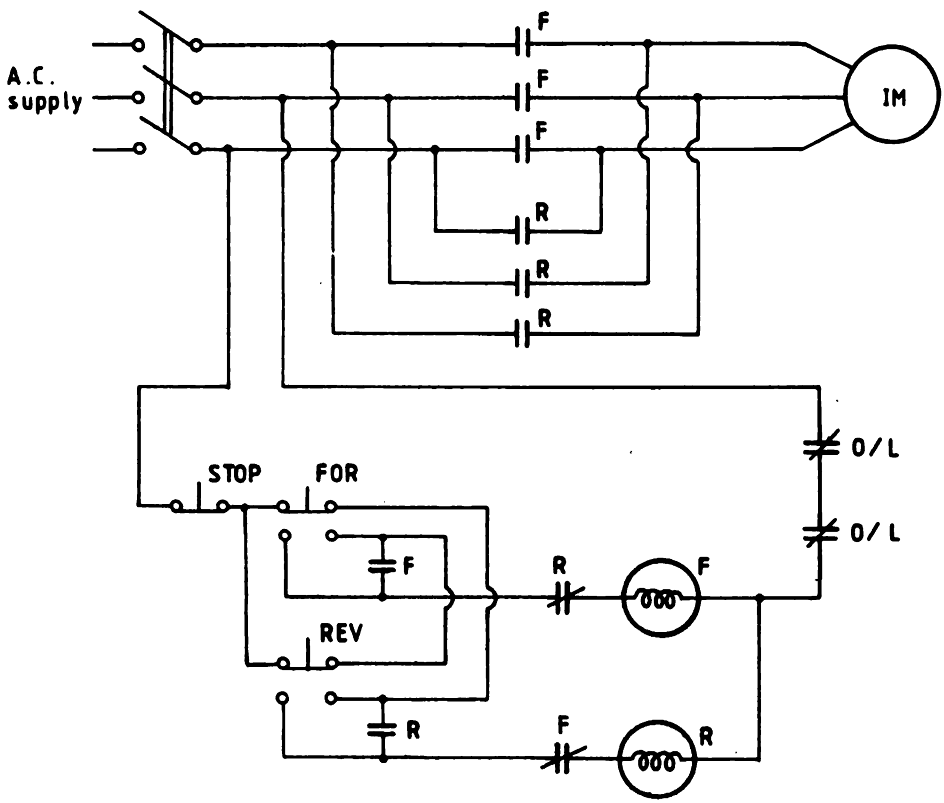


Fig. 7.6 Direct-on-line (DOL) starter for induction motors with provision for speed reversal.

induction motor, which rotates in the forward direction. When the rotation is required in the reverse direction, the reverse push button REV is pressed. The reverse contactors R pick up giving supply to the motor in the opposite phase sequence through R contactors. When the motor is to be stopped, the push button STOP is pressed.

7.3.2 Auto-transformer Starter

The auto-transformer starter (Fig. 7.7) is described. The link K_2 is directly connected to the common terminals of the contactors (U, V, W), Timer (T) and indicating lamp (IL). The circuit works on

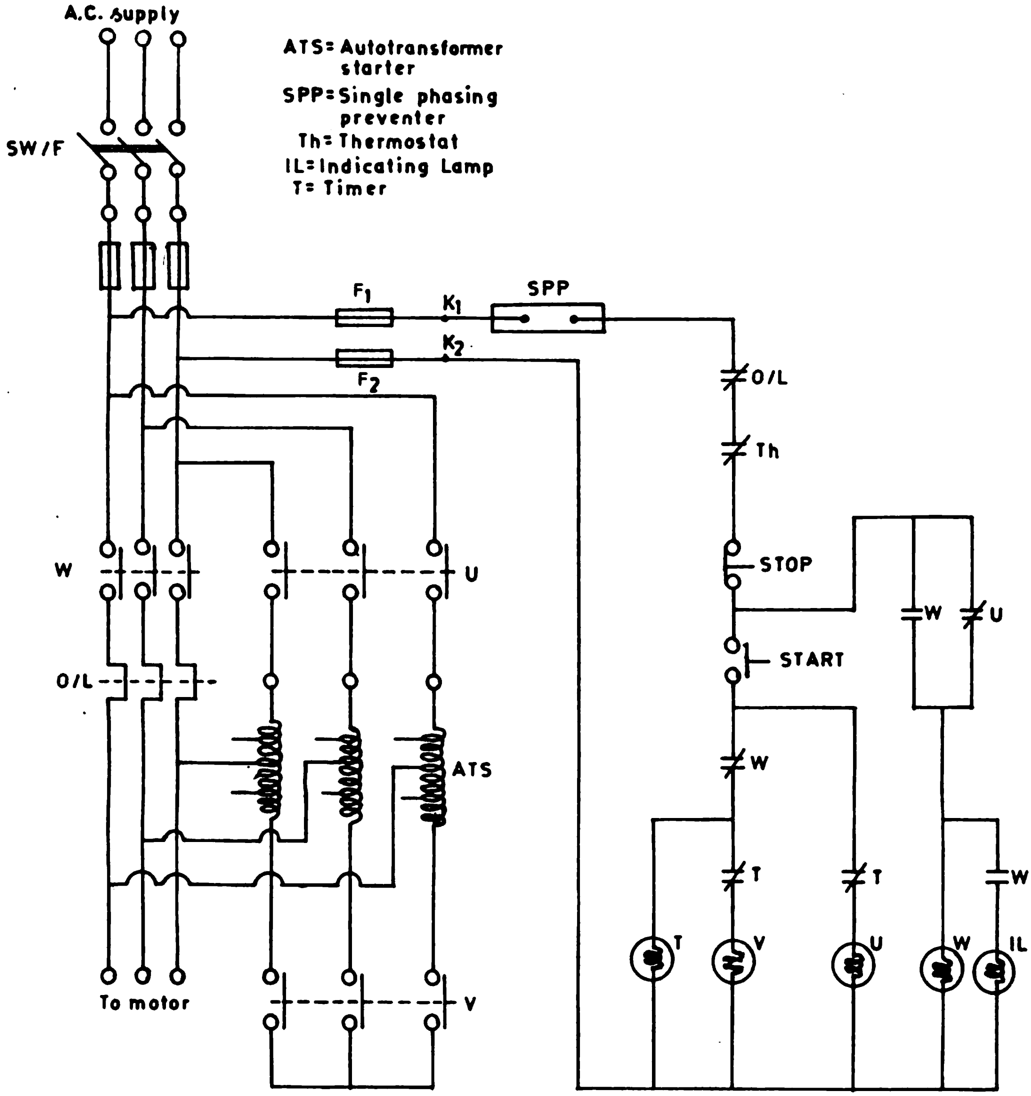


Fig. 7.7 Auto-transformer (Korndorfer) starter for induction motors.

415 V ac supply through control fuses F_1 and F_2. When the start push button is pressed, the timer T is energized through contact (NC) of W. The autotransformer is connected in star by the operation of contactor, V through the off delay contact (NC) of timer T, which closes instantaneously and then opens after the preset time delay. The contactor U is also energized through another off delay contact of timer T. When U is de-energized after the preset time delay of timer T, the contactor W is energized and the contact (NC) of U opens. The NO and NC contacts of W close and open respectively. The timer T is de-energized. The indicating lamp, IL glows. During the moment of closed condition of V and W, smooth transition from auto-transformer to direct-on-line (DOL) takes place. The start push button is locked through the contact (NO), W and also the contact (NC), U. Thus in continuous running condition only, the contactor W remains energized and the indicating lamp glows.

7.4 MASTER CONTROLLER FOR WOUND ROTOR INDUCTION MOTORS

The circuit of the master controller, shown in Fig. 7.8, for a wound rotor (slip ring) induction motor used in cranes, performs the functions of forward and reverse running, plugging and dynamic braking as well as emergency braking in case of power failure.

7.4.1 Starting

The control can be established by means of a master switch (Fig. 7.8). The crosses indicate the switch positions at which the particular contactors are energized. The master switch handle is notched from one position to another. When the master switch handle is notched from the neutral position to any of the forward positions, the contactor F picks up via 'd' line of the control circuit.

As a result, F contacts in the stator lines close and also the contactor M picks up through the contact (NC) of DB, closing M contacts in the stator lines, thus, connecting the stator terminals to the supply. The frequency sensing relays have been employed for starting, reversal and speed control. Over-current relays protect the motor against short circuit and thermal relays against overload. The plugging relay PR, which is a frequency sensing relay, is tuned to pick up at 51 Hz or so. The frequency sensing relays, 1FR, 2FR and 3FR are tuned to drop out at frequencies, say, 40, 20, and 10 Hz, respectively. Initially at starting ($\omega_m = 0$), the rotor frequency is 50 Hz. So, all relays, 1FR, 2FR and 3FR are energized with their contacts closed. Thus, the contactors, 1A, 2A and 3A are energized with the contacts, 1A, 2A and 3A open, which introduce maximum resistance in the rotor circuit as needed for starting. Only the plugging relay PR is de-energized with its contact PR open, thus the contactor P also remains de-energized with its contact (NC), P closed.

Following this, if the master switch handle is advanced to position no. 5 rapidly, the accelerating relays, 1FR, 2FR and 3FR de-energize successively, being governed by their respective frequency settings, as the motor accelerates and its speed increases resulting in the reduction of the rotor frequency. The accelerating contactors, 1A, 2A and 3A thus short circuit the steps of the resistance in the rotor circuit. The motor operates on natural characteristic after the contactor 3A closes.

The motor can be started only from off position. The under-voltage relay UV disconnects the control circuit in case of under-voltage including power failure. If the master switch has been notched in the 3rd, 4th and 5th position, the motor operates on the corresponding rheostatic characteristic. As the handle of the master switch is shifted to the neutral position. all contactors are de-energized.

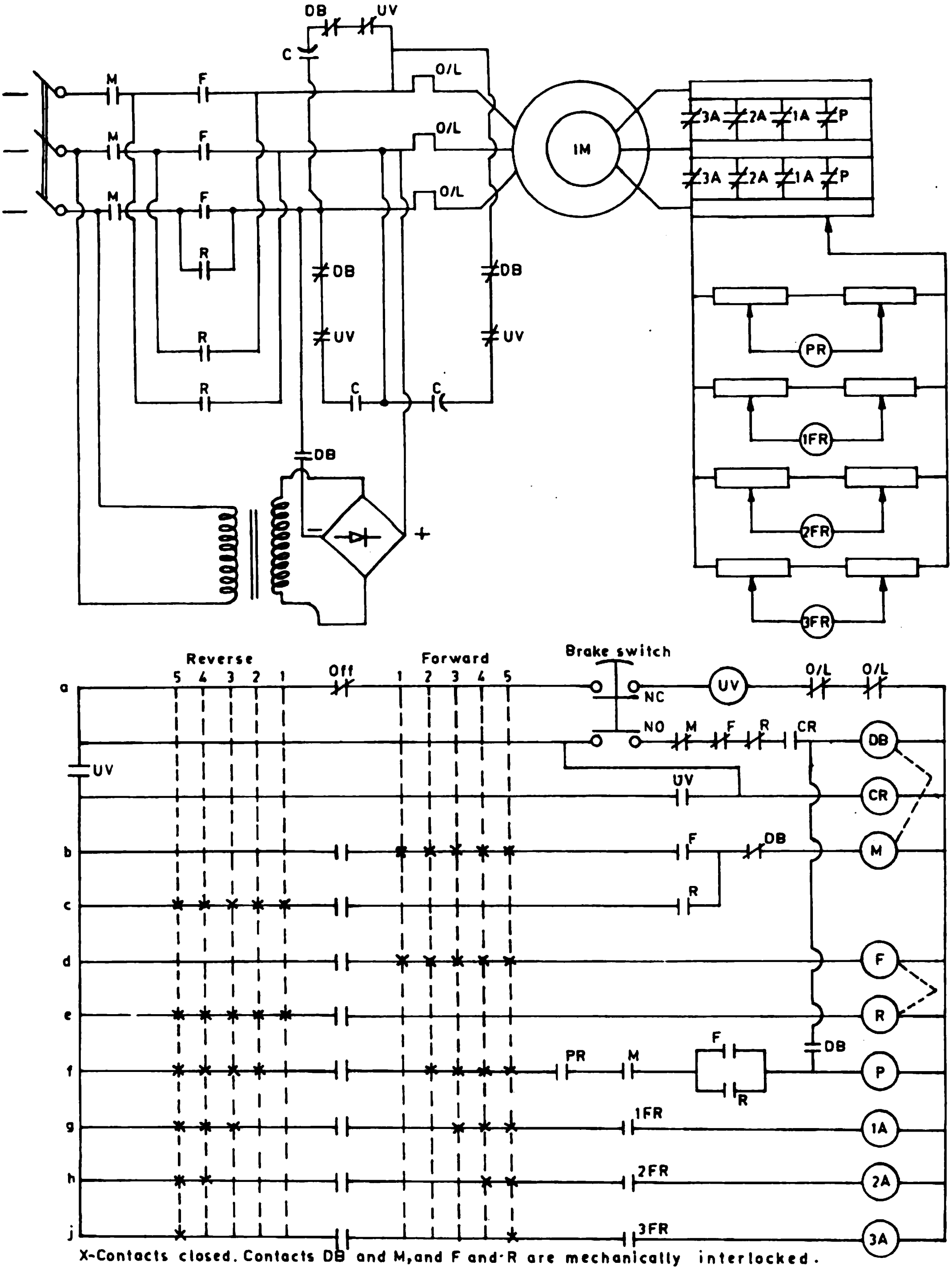


Fig. 7.8 Master controller for wound rotor induction motors.

When the handle is further shifted to the reverse position, the reverse contacts in the supply lines close. As a result the rotor frequency is more than the supply frequency, if the motor runs in the forward direction. As the rotor frequency exceeds 50 Hz, the relay PR is energized, thus energizing the contactor P when its contacts open, which introduces extra resistance in the rotor circuit. The plugging relay PR drops out immediately after the end of the transient process at the preset frequency of 51 Hz, when the motor speed is nearly zero (standstill). The contactor P drops out, thus, short circuiting the plugging section of the rotor resistance through its contacts, now getting closed. The mechanical interlocking arrangements have been provided between (a) F and R, and (b) DB and M. The relay CR performs the re-switching of the control circuit in a proper manner for dynamic braking.

7.4.2 Plugging and Speed Reversal

To perform the plugging and reversing operations of the drive, the master switch handle must be moved into one of the five reverse notch points. The handle is at first brought to the off position, when the supply is cut off from the stator. If the handle is now put in the fifth notch, for instance, the contactors in the control lines (c, d, e, f, g, h and k) of the master switch are energized. The contactor M picks up via 'c' line of the control circuit. The contactor R picks up via 'e' line, connecting the supply lines to the stator windings in opposite phase sequence, while the rotor still continues to rotate in the forward direction. The counter current braking (plugging) operation takes place. Initially, the rotor frequency is almost double the supply frequency. All the frequency relays including the plugging relay PR are energized. The rotor slows down with full resistance in the rotor circuit. When the rotor speed reaches close to standstill (51 Hz), the frequency relay PR drops out. Consequently, the plugging resistance is short circuited. If at this instant, the handle of the master switch is moved to off positon, the motor comes to a final stop after some time. If the handle of the master switch is left in this position, acceleration in reverse direction takes place, cutting out the rotor resistance step-by-step.

7.4.3 Dynamic Braking

Dynamic braking can be achieved at any working point of the master switch. To perform dynamic braking, the footswitch should be pressed and kept in this position. Immediately, the UV relay, but not the CR relay, drops out. Consequently, the left-side UV contact opens disconnecting the control circuit from the supply, i.e. M, F or R, P, 1A, 2A and 3A contactors drop out disconnecting the stator from supply. Then, the contacts (NC), M and F or R close with the CR contact also remaining closed. Consequently, DB contactor picks up closing its contacts with the rectifier output voltage appearing across the two terminals of the stator. The third DB contact in the vertical line of the control circuit closes to energize the P contactor. The latter picks up and opens its contacts in the rotor circuit, thus introducing the plugging set of rotor resistance for dynamic braking.

If the supply happens to fail, the UV relay is immediately de-energized cutting out the whole control circuit. The UV contact in the capacitor bank circuit is closed. As DB contactor remains de-energized, DB contacts are also closed at that time. Then, the capacitor bank is connected to the stator lines, achieving dynamic braking (AC braking) of the induction motor with self-excitation.

7.5 STARTERS FOR TWO-SPEED POLE CHANGING INDUCTION MOTORS

7.5.1 Starter for Two-Winding Two-Speed Pole Changing Induction Motors with Provision for Reversal at Both Speeds

The starter circuit for the motor is shown in Fig. 7.9. A selector switch in association with a pair of reversing contactors, F and R, sets the desired direction of rotation. Low and high contactors

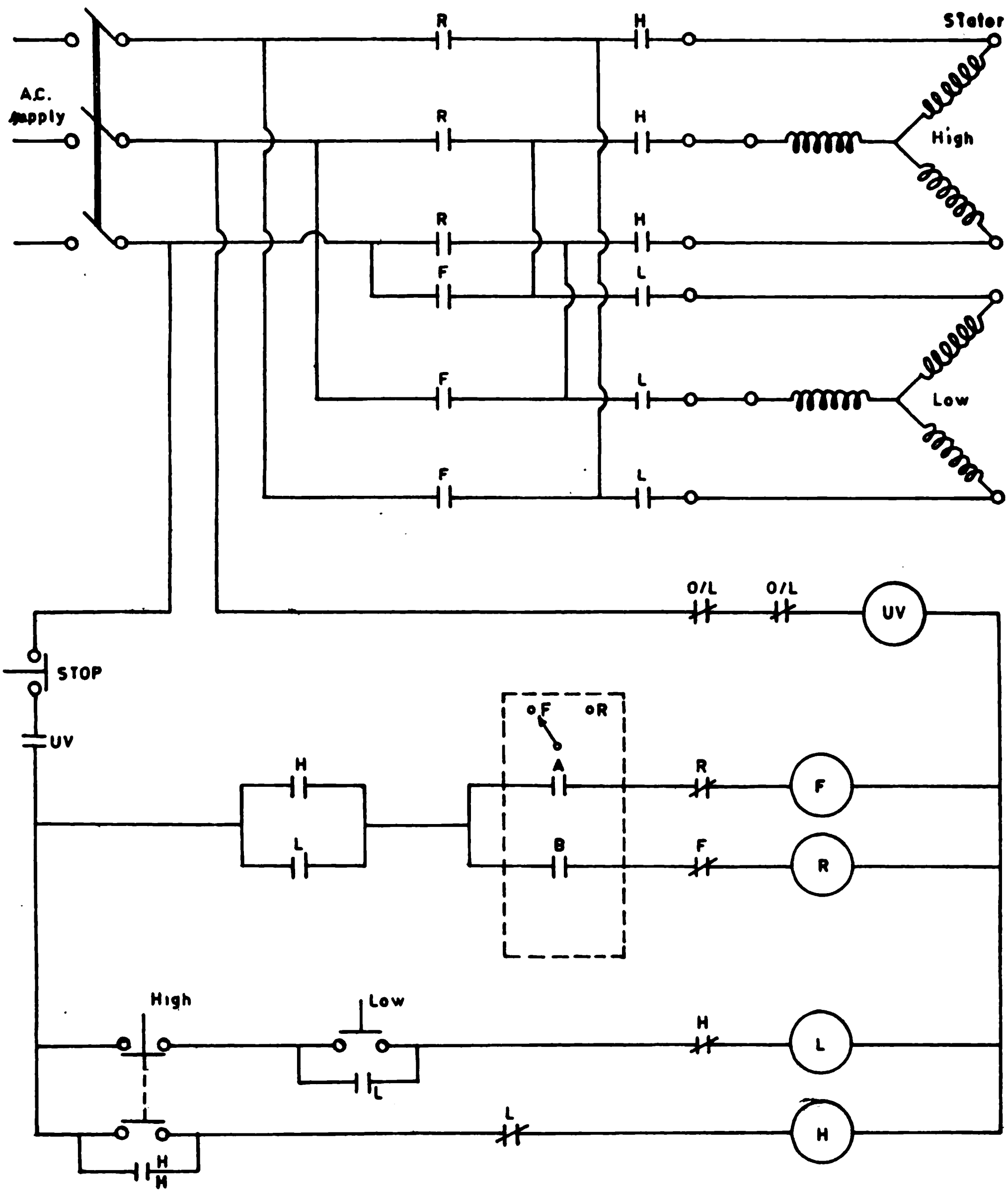


Fig. 7.9 Starter for two-winding two-speed pole changing induction motors with provision for reversal at both speeds.

operate, when the proper push button is pressed. Thus, if the selector switch is set at F to close the contact A and open the contact B, only the forward contactor F is energized. If the low push button is pressed, contactor L is energized. If the high push button is pressed, contactor H is energized. Similarly, with the selector switch set at R to open the contact, A, and close the contact B, only the reverse contactor R is energized. Then again, either of the push buttons, L or H, can be pressed to run the motor at the speed as desired.

7.5.2 Starter for Single-Winding Two-Speed Pole Changing Induction Motors with Provision for Speed Reversal (Constant Torque)

The starter circuit for a pole changing induction motor, which can be reversed on high speed, is shown in Fig. 7.10. The circuit is developed for constant torque operation. Table 7.1 shows the connections for low/high speed operations for Fig. 7.10a.

Table 7.1

Speed	*Supply to*			*Terminal connections*
	L_1	L_2	L_3	
Low	T_1	T_2	T_3	T_4, T_5, T_6—open
High	T_4	T_5	T_6	T_1, T_2, T_3—together

The motor can run either in forward or in reverse direction at high speed. At low speed, the motor can run in forward direction only. The motor has to start from rest for HF, HR or LF operations. If the push button LF is pressed, the control relay 1CR is energized closing its contacts, which energizes the coil LF. The contacts LF in the supply lines are then closed. The contacts, HF and HR remain open. Similarly, HF or HR operation takes place, if HF or HR push button is pressed.

7.5.3 Starter for Single-Winding Two-Speed Pole Changing Induction Motors with Provision for Speed Reversal (Constant Horsepower)

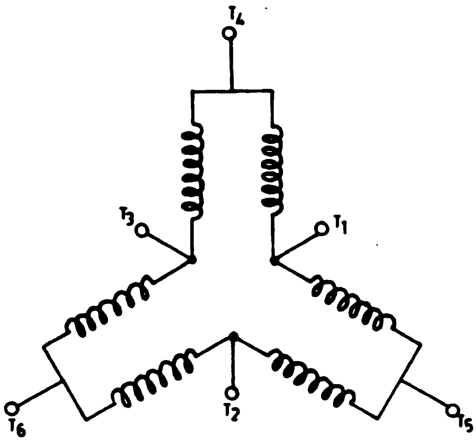
The starter circuit shown in Fig. 7.11 is used for constant horsepower operation of pole changing induction motors run in either direction, forward or reverse. Table 7.2 shows the connections for low/high speed operations for Fig. 7.11a.

Table 7.2

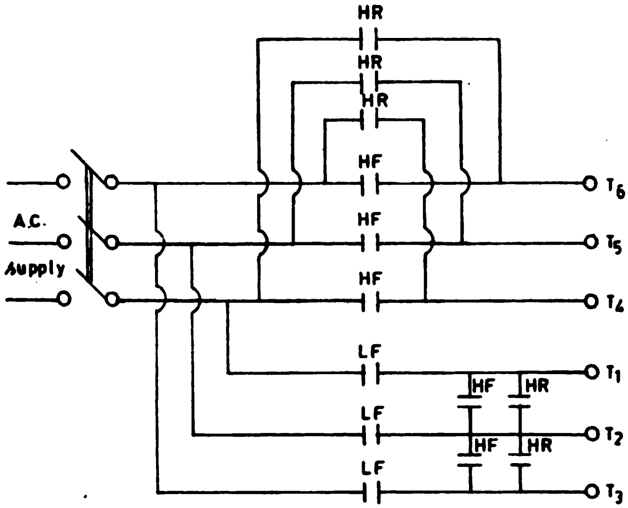
Speed	*Supply to*			*Terminal connections*
	L_1	L_2	L_3	
Low	T_1	T_2	T_3	T_4, T_5, T_6—together
High	T_4	T_5	T_6	T_1, T_2, T_3—open

Interlocking contacts have been included in the circuit to provide the following functions:

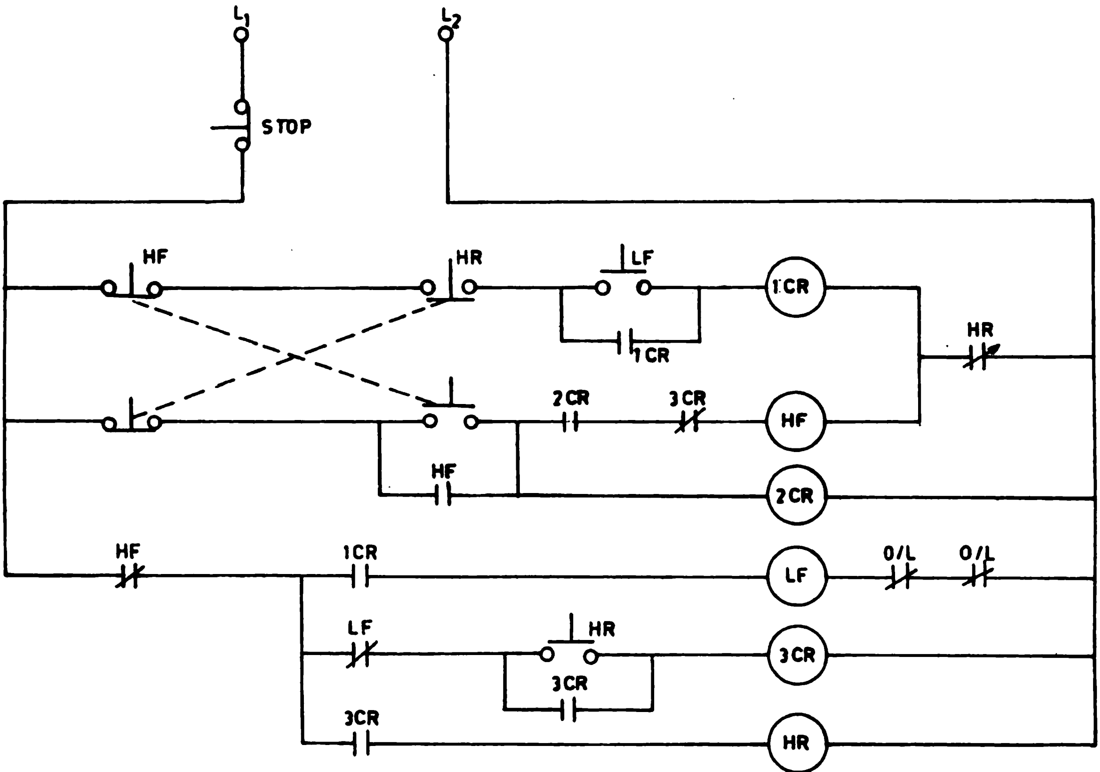
1. The motor must be started in low forward or low reverse direction. It cannot be started in high forward or high reverse direction. For this purpose the contacts, CR-TO (Control Relay Time Operated) are open in the HF and HR contactor circuits.



(a) Terminal markings.

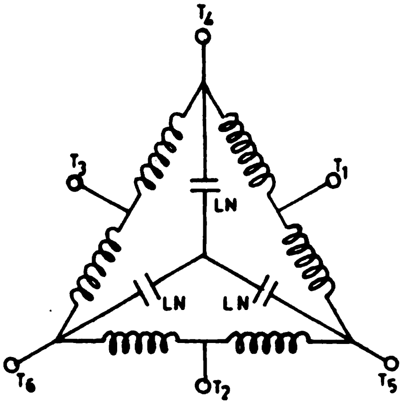


(b) Terminal connections to supply via contacts.

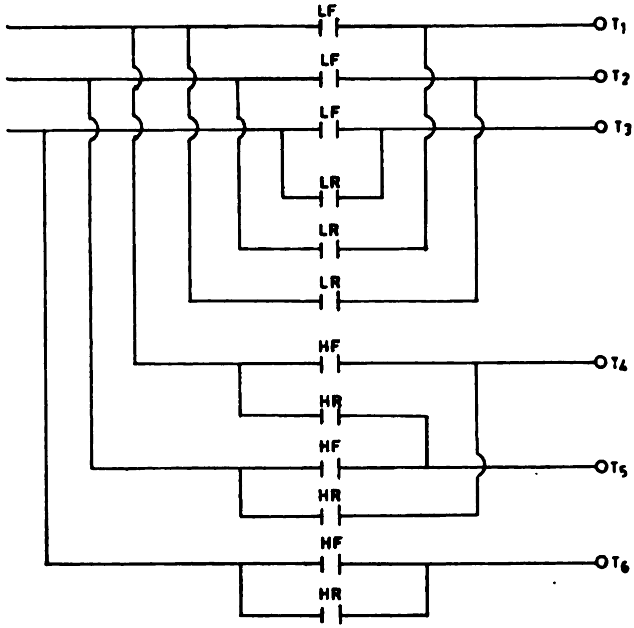


(c) Starter circuit.

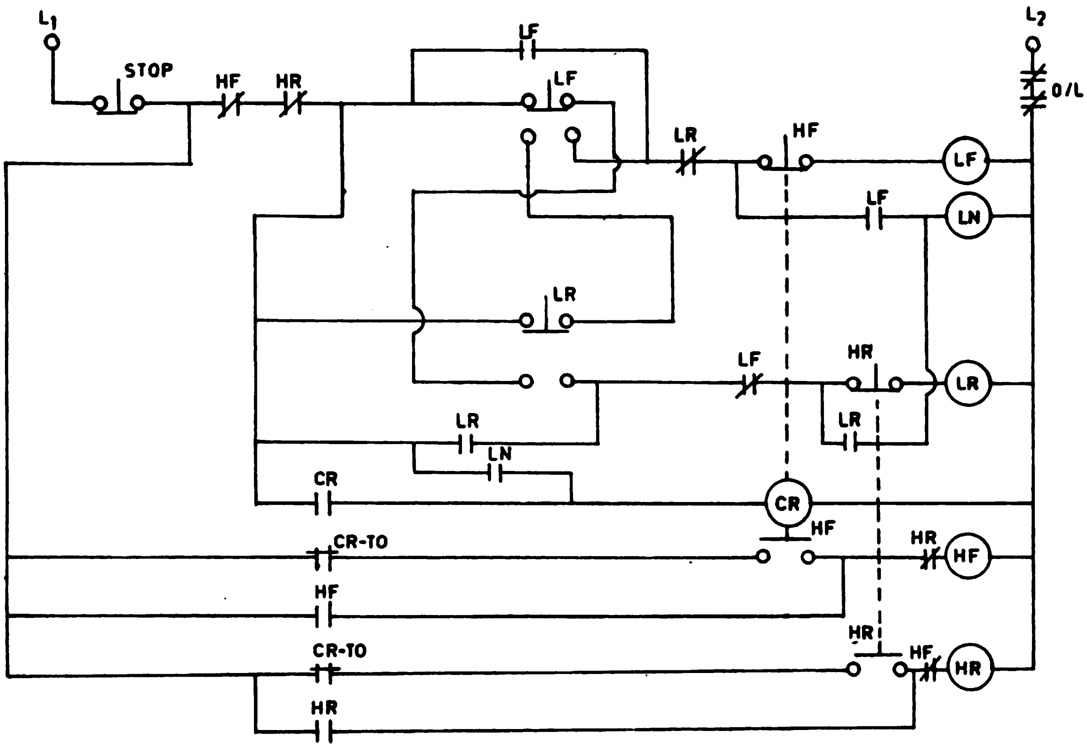
Fig. 7.10 Starter for single-winding two-speed pole changing induction motors with provision for speed reversal (constant torque).



(a) Terminal markings.



(b) Terminal connections to supply via contacts.



(c) Starter circuit

Fig. 7.11 Stater for single-winding two-speed pole changing induction motors with provision for speed reversal (constant horsepower).

2. A shift can be made from low forward to high forward or from low reverse to high reverse after a time delay set by the control relay CR. The two CR-TO contacts close, when the CR relay delay time runs out. Thus, the high forward or high reverse push button remains ineffective until these contacts close. This prevents the motor to shift to high speed until the motor has accelerated fully on low speed.

3. To shift from low forward to low reverse and vice versa, it is necessary to go through the stop sequence.

4. To shift from high forward to low forward, or from high reverse to low reverse, it is necessary to pass through the stop sequence. This is because HF(NC).or HR(NC) interlock (both in series) in LF control circuit is open. The LF or LR push button is ineffective until the HF or HR contactor drops out.

7.6 PROGRAMMABLE LOGIC CONTROLLER (PLC)

A programmable logic controller (PLC) is, nowadays, a key component for industrial automation. The whole range of logic for sequential control of equipment can be implemented by software, which is stored in the memory of PLC and retrieved by the microprocessor (CPU), whenever control is required. A PLC, therefore, acts as an open loop controller.

A number of ladder diagrams have been discussed earlier in connection with starting and braking of electric drives. In hardware logic control circuits, the individual logic blocks are connected by wires. Hardware logic control is dedicated to a particular (sequential) control of equipment. It is very difficult to change with hardware controllers even a single sequence without loss of production. Hence, hardware sequential control is not at all flexible. In contrast, PLCs are more reliable and flexible. The program stored in the memory can be readily changed (modified). Hardware control is gradually being replaced by software control as PLC hardware is fully standardized with fault diagnostics developed for this system. The novelty of PLC is that it can operate in industrial environment with temperature ranging from 0 to 60°C, and humidity ranging from 0 to 90%.

The PLC is a programmable device, which executes functions such as logic, timing, counting, arithmetic operations and data manipulations. It constantly monitors the status of the input devices, processes the input data as programmed, makes decisions and modifies the outputs accordingly. It interfaces with the input and output devices as shown in Fig. 7.12, which also gives the basic components of a standard PLC system.

A control program consists of a series of instructions which are executed by the CPU of PLC for sequential control of a system. To develop a program, the ladder diagram discussed earlier can be used as an aid. For detailed programming method, the operation manual of the manufacturer may be used.

The instructions are entered from the keyboard with the processor in the **Program** mode. The program may be edited in the **Adjust** mode and executed in the **Run** mode.

There are various types of instructions, such as In/Out, Relay type, Timer/Counter including data manipulation and other types of instructions. The program is written by using user's symbols. A couple of simple ladder diagram programs are given as examples (Fig. 7.13).

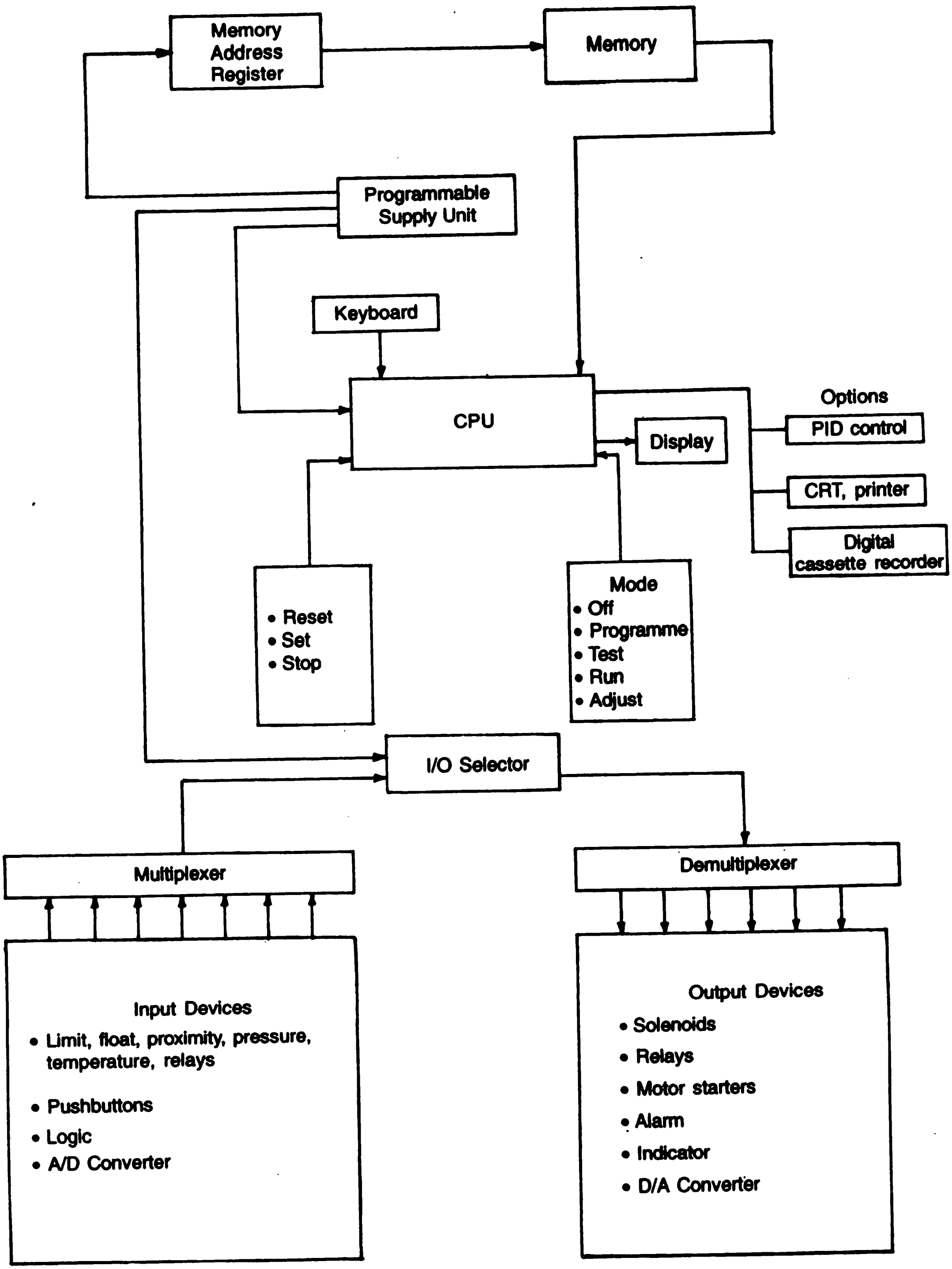


Fig. 7.12 Architecture of a programmable logic controller (PLC).

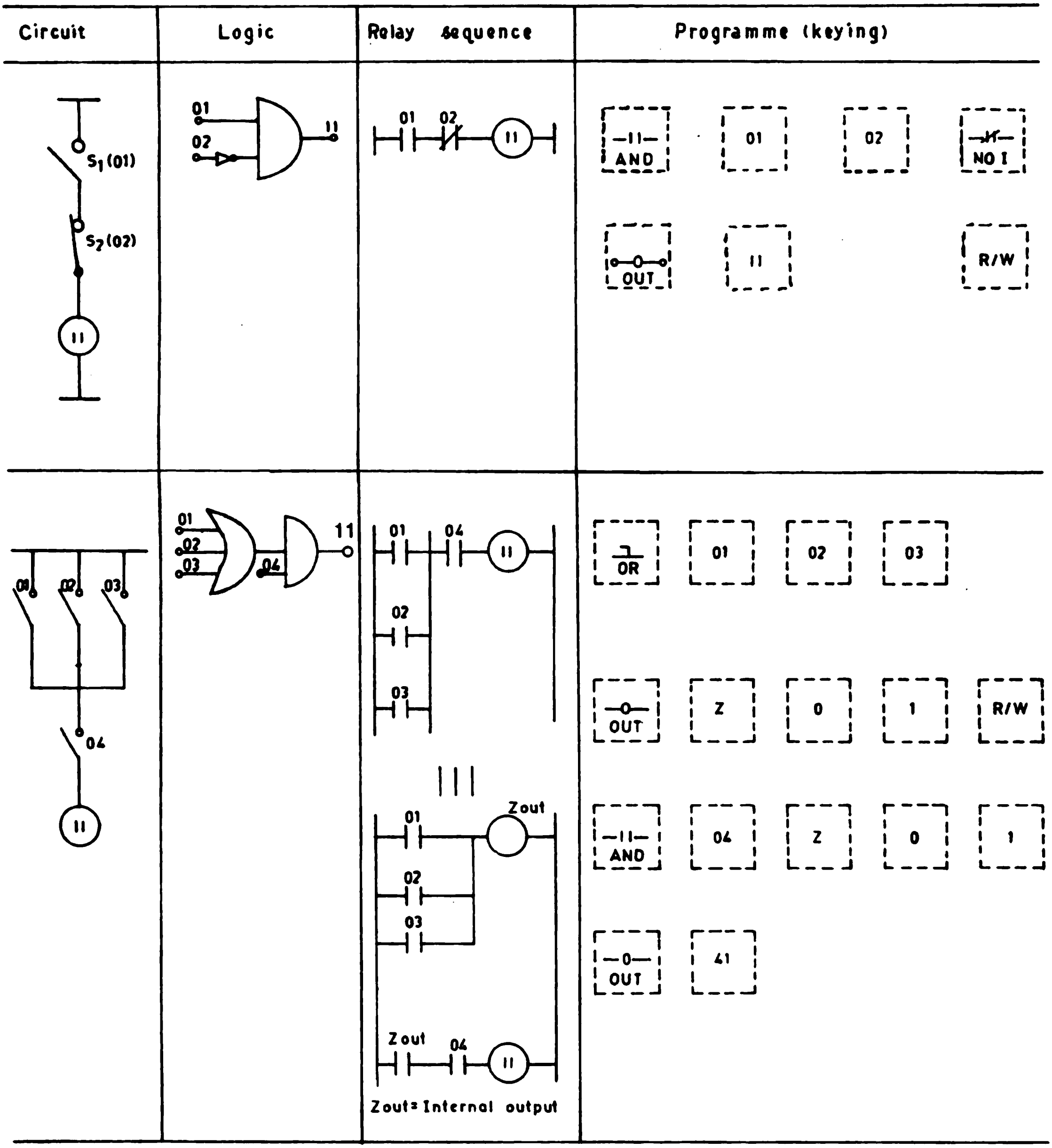


Fig. 7.13 Simple ladder diagrams with programs for PLC.

CHAPTER 8

INDUSTRIAL APPLICATIONS

8.1 INTRODUCTION

In the preceding chapters, the basic principles of electric drives have been presented along with their applications. In this chapter, the drive systems as used in different types of industries are discussed. There are certain common aspects of such drive systems, which are described initially in continuation of what has already been told about the four main types of drive motors and their control.

8.1.1 Synchronous Motor Drives

Advantages

1. Ability to work as synchronous condenser for power factor improvement with better power factor at low speeds.
2. Higher efficiency than that of an induction motor of comparable size.
3. Lower cost.
4. Maximum torque varies directly instead of square of voltage.

Disadvantages

1. Starting problems are faced.
2. A damper winding is required. If high starting and also high pull-in torque are required, a double cage damper winding is used.

Normally a synchronous motor drive has to produce starting torque of 150–200%, pull-in (synchronizing) torque of 110% and pull-out torque of about 175–180%. Synchronous motors are normally constructed in large sizes.

Motor-generator sets, air compressors, pumps and other constant speed equipment are examples of synchronous motor drives operating as synchronous condensers.

8.1.2 Induction Motor (Cage) Drives

A cage induction motor is very rugged and sturdy. It has the lowest initial cost per installation. Its efficiency and power factor are poor compared to a synchronous induction motor.

Normally direct-on-line (DOL) starters are used for motors with double-cage or deep-bar rotors. In industry, cage induction motors have maximum applications, and are used wherever speed control is not required. They are constructed in various sizes ranging from small to very large. Therefore, they have a very wide range of applications in industry such as water pumps, compressors, fans and line shafts, to name a few. Pole-changing induction motors may be used wherever two-speed operation is required. However, they cost 30–40% more than standard cage induction motors.

The development of PWM inverters for speed control of the induction motor operating at constant Volts/Hz ratio, has kindled hope for replacement of dc drives by ac drives in the near future. The disadvantage of this type lies in poor control at low speeds because of stator winding resistance dominating over leakage reactance.

8.1.3 Induction Motor (Wound Rotor) Drives

Various speed control methods are available for slip-ring induction motors as discussed in Chapter 4. The efficiency and power factor decrease with decrease of speed. The cost is also higher. However, the operation of a slip-ring induction motor incorporating an automatic slip regulator and a flywheel for intermittent duty is so unique that it is worthwhile to discuss.

The load sharing between the flywheel and the motor without the automatic slip regulator has already been discussed in Section 5.6. The load sharing depends upon the speed drop and hence the rotor resistance. If a high value of rotor resistance is permanently used, the final speed is reduced resulting in production loss and also energy loss in the rotor circuit. An automatic slip regulator overcomes the above difficulty. There are two types of automatic slip regulators: liquid and contactor. The scheme of liquid type automatic liquid slip regulator as shown in Fig. 8.1 is used in rolling mills, electric hammers, presses and such other equipments.

The required resistance is automatically inserted into the rotor circuit as the load is increased. It is assumed that initially the motor is operating at no-load at its minimum resistance position. As the load appears on the main induction motor shaft, its torque increases following the natural characteristic (Fig. 8.2) along *Oa* until the automatic slip regulator comes into action at a specified value of the torque equal to KT_{nom} ($K > 1$), which depends upon the counterweight. The torque increases exponentially from *O* to *a*. During the period of its action from the point *a*, the control motor operates in the direction of the counterweight raising the moving electrode.

The operation of the main induction motor shifts from one speed-torque curve to another maintaining constant torque along *ab* so long the load is there on the shaft up to the point *b*. During the period from *a* to *b*, the energy is supplied by the flywheel to the load. When the motor is unloaded, the current and the torque of the control motor decrease, the moving electrode goes down, the speed increases maintaining constant torque along *ba*. On the load diagram the torque remains constant up to the point *c* during the no-load period, when the motor is transferred to the natural characteristic at the point *a*. The action of the automatic slip regulator stops at this instant. The torque is further decreased along the natural characteristic, thus increasing the speed. During the off period, the energy is stored in the flywheel. The torque decreases exponentially along *cd* on the load diagram. Motor load is thus reduced and smoothened out. In this way it is possible to select a motor of 30% to 50% capacity of maximum load requirements. The rotor resistance can be controlled in this method by using contactors.

Wound rotor induction motors are used for flywheel equipped motor-generator sets working under shock loading conditions. These motors are used for mine winder drives. They are also used for line shafts, pumps, lifts, cranes, haulage, etc. They are used in paper making machines, textile

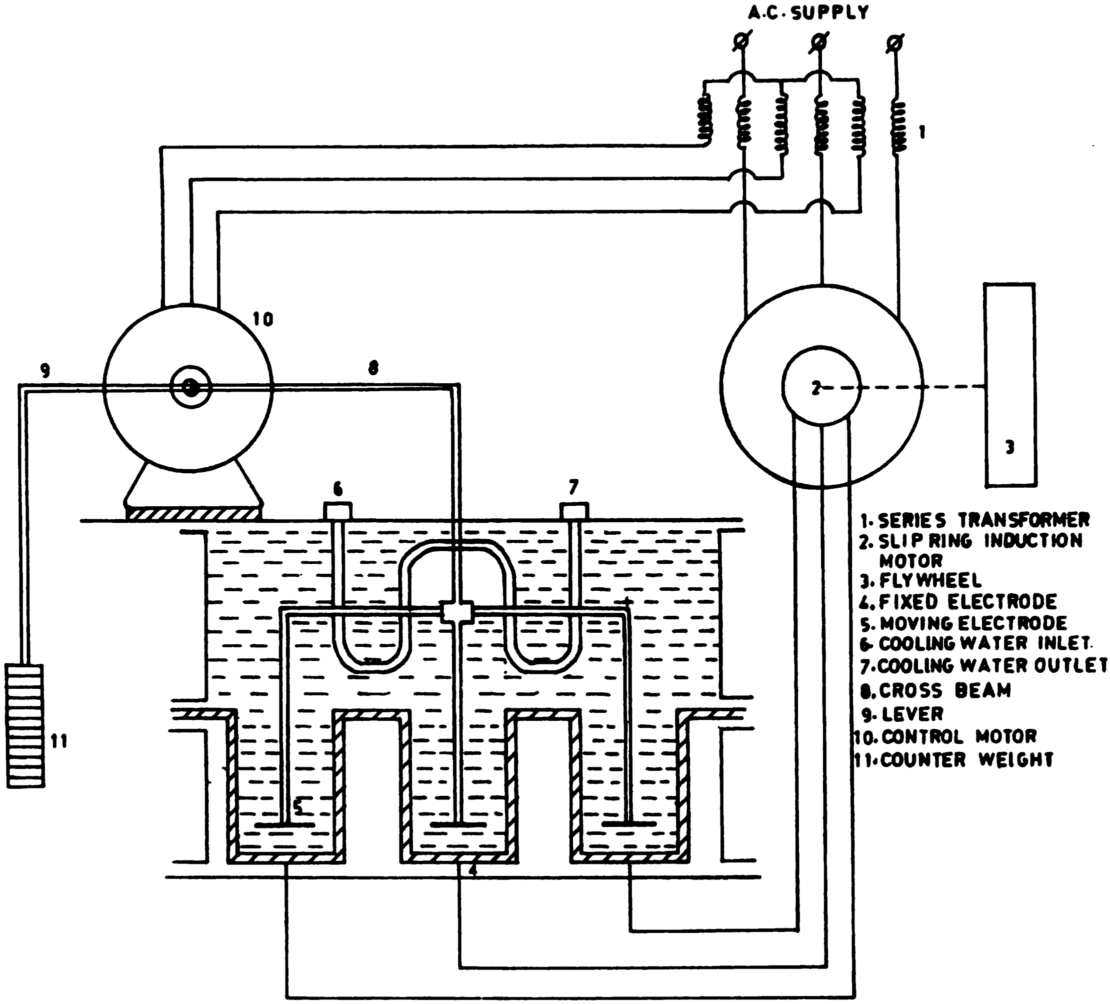


Fig. 8.1 Liquid type automatic slip regulator.

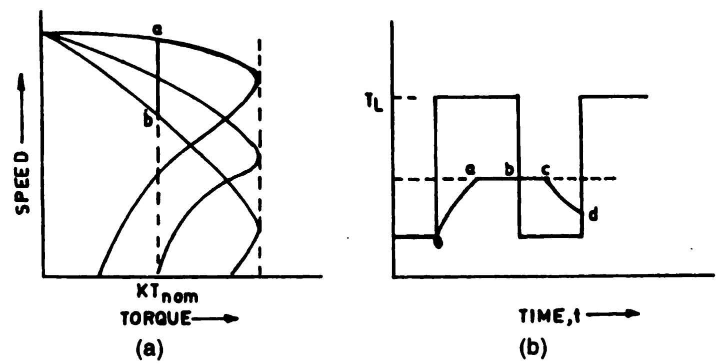


Fig. 8.2 (a) Speed-torque characteristics of induction motors; and (b) Load diagram under the action of an automatic slip regulator.

mills, printing press and other industries. Reciprocating pumps, crushers, shearer and lump breakers in coal mines and ball mills in cement factories require high starting torque, which may be two to three times the full-load torque due to high inertia and friction at standstill. Full-load torque is required for normal running. Slip-ring induction motors are satisfactory for such operations.

8.1.4 DC Motor Drives

Whenever there is a demand for a wide range of speed control in either direction (reversible drives) with better regulation and simplified mode of control, the choice falls on dc drives, for example, rolling mill drives, paper mill drives, etc. DC motors have the advantage of ease of operation at different speeds. The operating characteristics can be varied. DC series motors are used in cranes/ hoists. They are also ideal as traction motors.

But dc motors are most expensive of all. The dc shunt motors with constant speed characteristics are used in rolling mills, paper mills, etc. where precise speed control is necessary. The Ward-Leonard method of speed control incorporating continuous feedback permits smoother operation and greater accuracy.

8.1.5 Control Systems

Modern control systems are designed to increase the production time and reduce the down time, with a view to achieving the highest possible efficiency to cope with the increasing competitive market. There has been revolutionary progress in the field of electric drives. Till 1970, the magnetic amplifier and amplidyne were used extensively for motor control. The development of transistors, analog and digital chips (ICs), thyristors and allied devices, bipolar transistors (power), MOSFETs and microprocessors including computers has resulted in improvement of controllers for electric drives.

Electronic control may be of two types: (a) analog (linear) control, and (b) digital control. Analog (linear) control is used, when good accuracy, reliability and speed of response are needed. The highest degree of the above qualities of control is obtained using digital control. The reliability means to fulfil all predefined conditions of control within the preset time period.

In conventional control, relays and contactors are used in most places. The programmable logic controller (PLC) has brought about modernization replacing the conventional relay and contactor systems. The PLCs are being used for performing complex automation tasks at very low cost with inherent flexibility and quick commissioning. These are microprocessor-based controllers. Nowadays, P, PI and PID controllers are also available.

P-controller. This has very fast response because of the absence of time constant. Input and output voltages are proportional. It is very difficult to have the output equal to the reference or set value.

I-controller. In this controller, the output voltage is time integral of the input voltage. The response is slow. Time constant and integrating elements are incorporated.

D-controller. The response of an ideal D-controller is instantaneous. It is not possible to make an ideal D-controller. Normally this is used along with an I-controller.

PI-controller. It has the properties of both P and I controllers with the advantage that the steady-state error is zero such that the output is equal to the reference or set value.

PD-controller. It is almost impossible to realise an ideal PD controller, which has the properties of both P and D controllers.

PID-controller. This is an embodiment of all the features of P, I and D controllers and can be tuned to meet the transient response of the controlled system as desired with zero steady state error as in the PI-controller.

Microprocessor-based controllers improve the dynamic response of the plant and are now used extensively in different industries. Automation tasks such as logic control, closed loop control, PID control, computing, condition monitoring, fault diagnostics, signalling and data acquisition/storage are achieved with the help of processor-based systems. Process control is mostly effected through feedback control. Multi-loop controllers, being processor-based systems used in closed loop control, are very simple to handle and have fast speed of response.

Computer control is called for when the following process conditions prevail in the industry:

(a) To obtain high production rate in repetitive work, which causes fatigue to the operator.

(b) When complex computations and decision making are involved, which are beyond the capability of the operator.

(c) When the operator cannot perform complex computations and make decisions in a specified time.

(d) When a large amount of data is to be handled.

With the above description of the common features of electric drives, attention is now devoted to applications of drives in different industries.

8.2 ROLLING MILL DRIVES

Rolling mills are mainly of two types: (i) primary reversing mills, and (ii) continuous mills. Blooming, slabbing and plate mills are the most important reversing mills, which reduce large ingots to smaller size blooms, slabs and plates.

8.2.1 Reversing Mill Drives

The primary reversing mills normally comprise slow speed drives, namely 40/80 rpm. The function of a blooming mill is to convert steel ingots into semifinished products. DC motors are used. These motors are directly connected over long spindles to the mill rolls. Voltage control is applied at constant torque up to base speed, and for speed control above base speed field control is applied, which represents constant horsepower operation. Blooming mill can be reversed from 50 rpm in forward direction to 50 rpm in reverse direction in just one second. The load diagram of rolling mills has already been discussed in Section 8.1.3 and also in Section 5.6. The primary mills have peak load for short durations. It is economical to use an Ilgner set in association with an automatic slip regulator for driving the mill motor of most reversing mills. The rolls run in both directions in sequence. The pass schedule depends upon the size of the ingot, its temperature, the capacity of the mill motor and the size of the final product. The motor capacity varies from 2.5 MW to 10 MW. When the power requirement of a reversing mill exceeds 5 MW, a twin drive is used—one to drive the top roll and the other to drive the bottom roll. The inertia is less for the twin drive. The twin drive does not require a pinion stand. Rolls of different diameters can be used with twin drives. Plugging and dynamic braking are incorporated in the control scheme of reversing mill drives. The selection of a proper mill motor depends upon the roll speed, reduction required, horsepower required and the degree of flexibility of the control.

8.2.2 Screw-down Mechanism

A screw-down mechanism is used to adjust the position of the top roll for each pass depending upon the reduction of the metal required. The screw moves axially when the nut is rotated, but its axial displacement is not there and the rotation of the screw is prevented (Fig. 8.3). The nut is with worm gear, which rotates in bearings inside its casing. The screw goes through the nut, but its rotation is prevented by a long keyway, the key being provided inside a guide for the screw. As the rotation is prevented it will move axially up and down according to the direction of rotation of the motor.

On primary mills, the screw-down operation is actuated by two dc compound motors of ratings 100–200 hp, mounted on the housing cap of the mill. For fast response, the inertia of the motors and the brake wheel should be low compared to the general purpose motors. Ward-Leonard control is applied. Amplidynes were used in old mills for torque feedback for automatic control of screw-down motors supplied by variable voltage motor-generator sets. A maximum roll lift of 1.6 m may

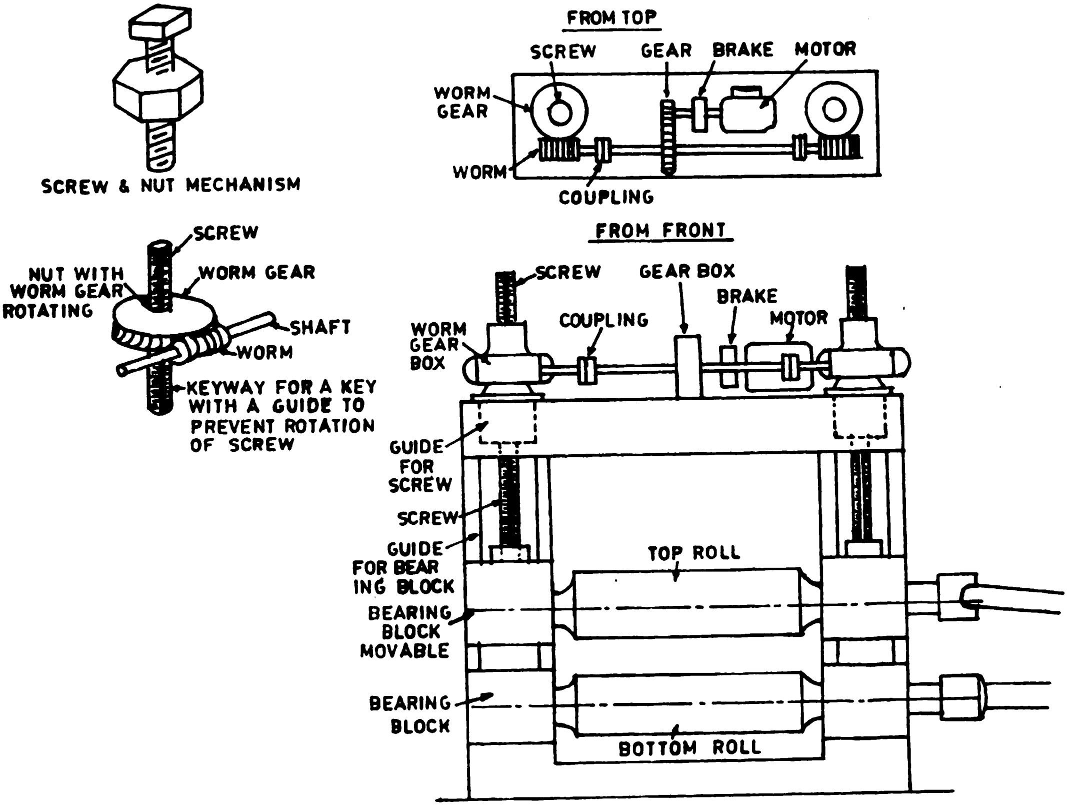


Fig. 8.3 Screw-down operation in a rolling mill.

be required with lifting speed up to 25 cm/s. Microprocessor-based thyristorized controllers may be used for speed control of screw-down motors for the entire pass schedule. The screw-down operation is not required in continuous mills or in three-high mills, where fixed passes are used.

8.2.3 Continuous Mills

The blooms are produced in a blooming mill. These are further processed to manufacture the finished products in secondary mills like billet, bar, structural mills, etc. There are various types of secondary mills: three-high, continuous tandem or zig-zag layouts. Structural mills produce a wide variety of products like angles, channels, rods and rails, etc.

In three-high mills, there are three rolls with two roll gaps in between providing rolling facility in either direction. A slow speed wound-rotor induction motor is used as a drive, which rotates continuously in one direction only. The flywheel is fitted with the rotor shaft of the motor, which is also controlled by an automatic slip regulator. In some mills, dc motors are also used to take the advantage of variable speed. The roll change is accomplished by tilting tables, which are also operated on the principle of screw-down mechanism. A slow speed slip-ring induction motor or a variable speed dc motor may be used for this purpose.

Continuous mills comprise a series of roughing stands followed by a finishing stand, all working in tandem. Although the reduction of metal takes place in all the roughing stands, the final thickness is obtained in the finishing stand. Modern continuous mills are provided with individual motor drives at each stand. Synchronous motors, wound-rotor induction motors or dc shunt motors are used through gear boxes. The rolling speed differs from stand to stand. The drives should run at coordinated speeds. The speed range lies between 400 to 600 rpm for continuous billet mills.

In continuous bar mills, numerous stands are used. The stands can be divided into a number of groups. Group drives are used. The speed range lies between 300 to 900 rpm.

Table rollers. Table rollers are driven either by individual dc motors or by a group of dc motors of suitable capacity through line shaft and gearing. The variable voltage method is applied to regulate the speed.

8.2.4 Cold Rolling Mills and Automatic Gauge Control

A cold rolling mill has a number of stands in tandem. More power is required for cold rolling than that for hot rolling. A twin drive is used. The top and bottom rolls of each stand are individually driven. Precise tension control is required, which is accomplished through speed control of roll drive motors. These motors are controlled by voltage variation with current feedback. The screw-down motors are voltage controlled as usual.

In Fig. 8.4, five stands have been shown of which the centre one serves as the reference with respect to which tension/speed is controlled. Thickness gauge is continuously monitored. The control scheme is implemented for an automatic gauge control. In case of over-gauge, the tension is increased. This function is performed by increasing the speeds of stands 4 and 5 and decreasing the speeds of stands 1 and 2. In case of under-gauge, the tension is decreased by increasing the speeds of stands 1 and 2 and decreasing the speeds of stands 4 and 5. Modern automatic gauge control is implemented by processor-based systems.

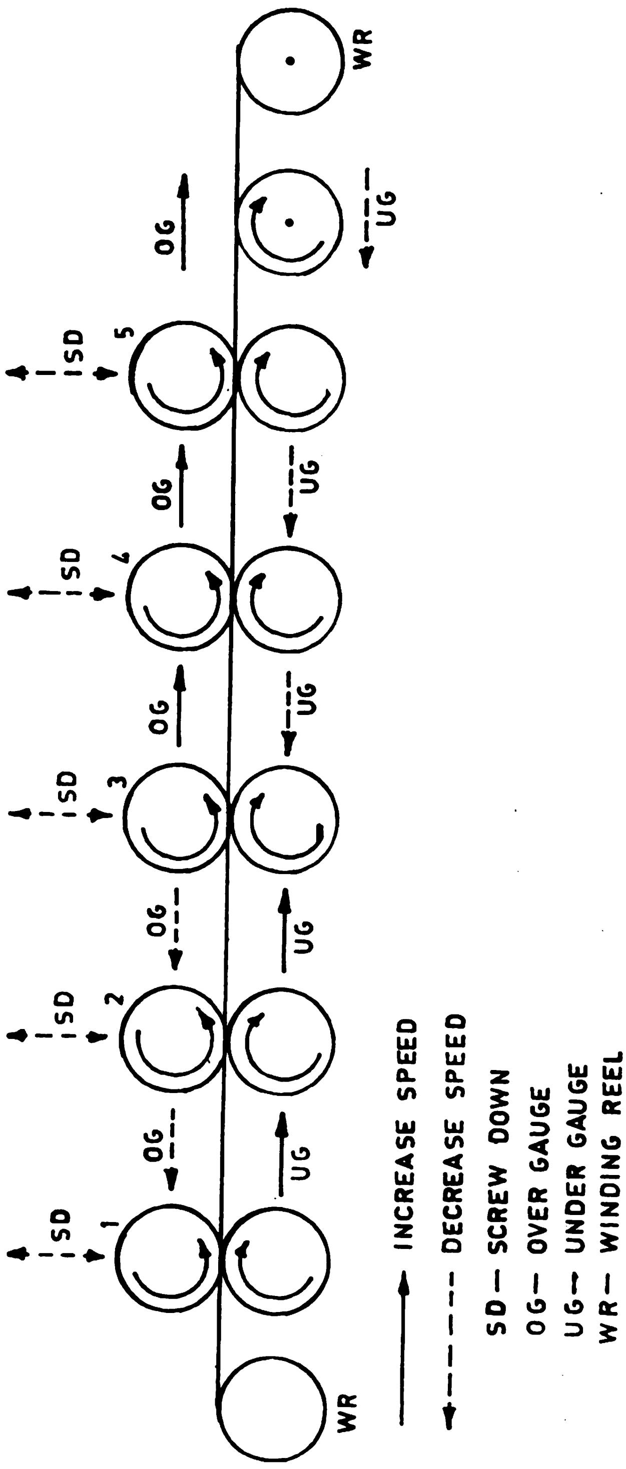


Fig. 8.4 Schematic representation of an automatic gauge control scheme in a cold rolling mill.

8.3 CEMENT MILL

A line diagram layout of a cement mill is shown in Fig. 8.5. The raw materials are fed to the mill in right proportion with the help of weighfeeders. A cement mill is a ball mill. The output of the ball mill is taken to a separator, which separates out the unacceptable coarse cement (less than 3000 blaine) and feeds it back to the mill. The fine acceptable product (above 3000 blaine) is fed to a cyclone which separates the air dust from the cement product. The cement is then transported to the silo.

8.3.1 Weighfeeder

In modern cement mills, a microprocessor-based control scheme incorporating a load cell using strain gauges is used to feed raw materials to the mill in the right proportion. The schematic diagram of such a controller is shown in Fig. 8.6. A separate drive motor (dc series motor) for each raw material is used to drive the feed table under which strain gauges are placed for measurement of weight of the material. The strain gauges are deformed by the load on the feed table and their resistance is consequently modified. The Wheatstone bridge is now unbalanced and its output voltage is fed back to the processor-based controller. The tachogenerator senses the motor speed.

$$\text{capacity}(\%) = \text{loading}(\%) \times \text{speed}(\%)$$

If the loading increases, then the speed decreases. This is automatically done by the processor-based controller. However, the weighfeeders can be operated in various modes, namely local, manual, interlock, etc. In local mode, start, stop and feed rates are fixed by the operator from a local control box installed near the operator. In manual mode, the feed rate is fixed from a remote control panel situated away from the operator, but start and stop operations are performed by the operator through the local control panel. Interlocked operations mean that the weighfeeder operation is fully remote controlled.

8.3.2 Mill Drive

A salient pole synchronous motor/synchronous induction motor is employed as a mill drive. A capacitor bank is used in case of a slip-ring induction motor for improvement of the power factor. The synchronous induction motor operates at 0.9 power factor (lead). The working voltage of a cement mill drive is commonly 6.6 kV. The starter used is of liquid resistance type. The speed of a cement mill is very low ranging from 15 to 20 rpm, which is achieved through a gearbox. Alternatively, the inverter-fed induction motor with constant Volt/Hz ratio can be used. But at low speed, cooling problems occur. The motor has to be, therefore, derated. However, the noise level is lower, harmonics and torque pulsations are eliminated, and motor efficiency and line power factor are uniformly high. Over-current and under-voltage relays are provided for protection. The starting current is restricted to 250% of full-load current and starting torque to 175% of full-load torque to avoid voltage dips. The pull-out torque is restricted to 240% of rated torque. The cement mill drives are normally designed for 2–3 consecutive starts per hour from cold condition and one start per hour during hot condition against full-load. They are also designed to operate at 50% overload for one minute, four times per hour at equal intervals.

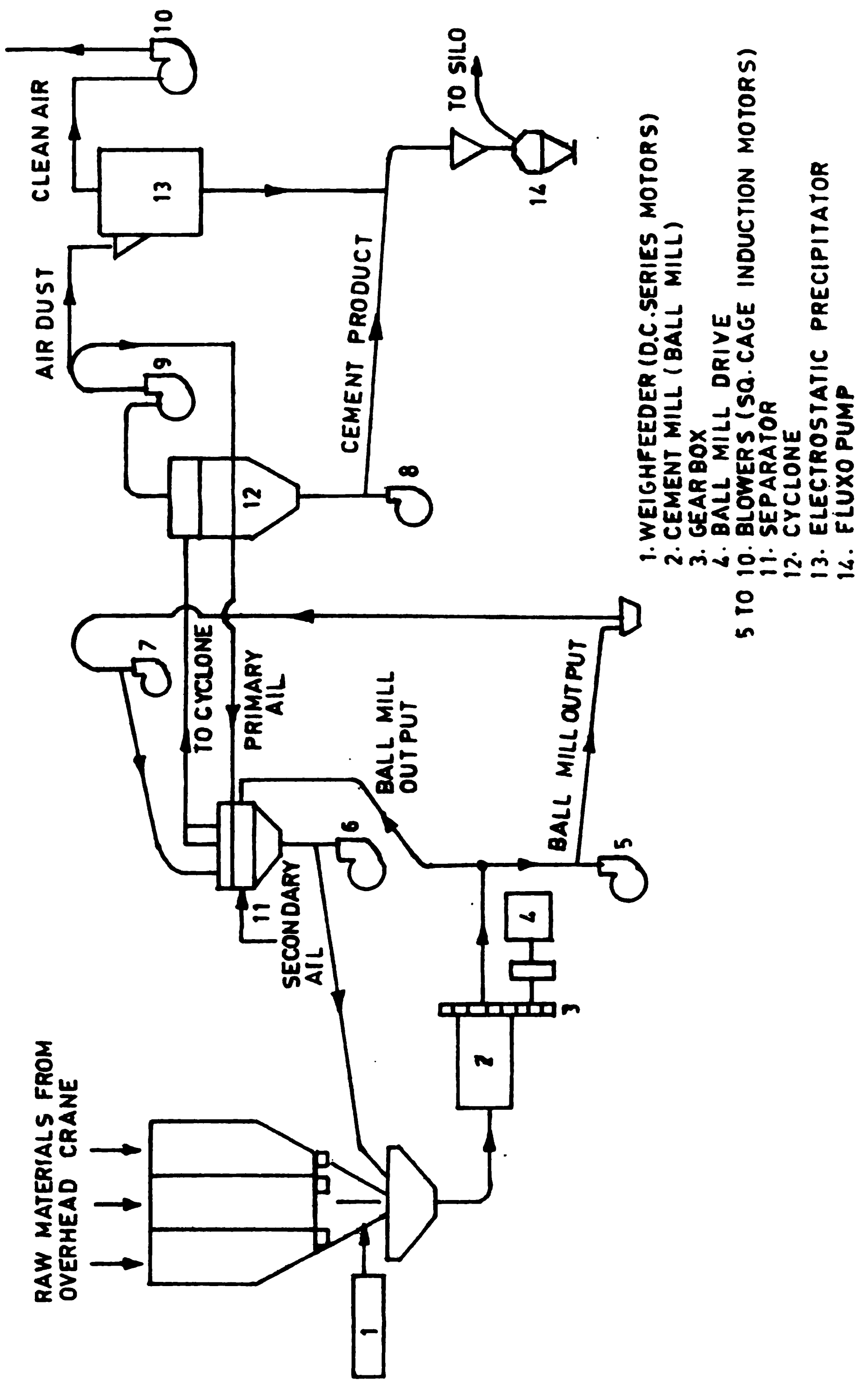


Fig. 8.5 Layout of a cement mill—Line diagram.

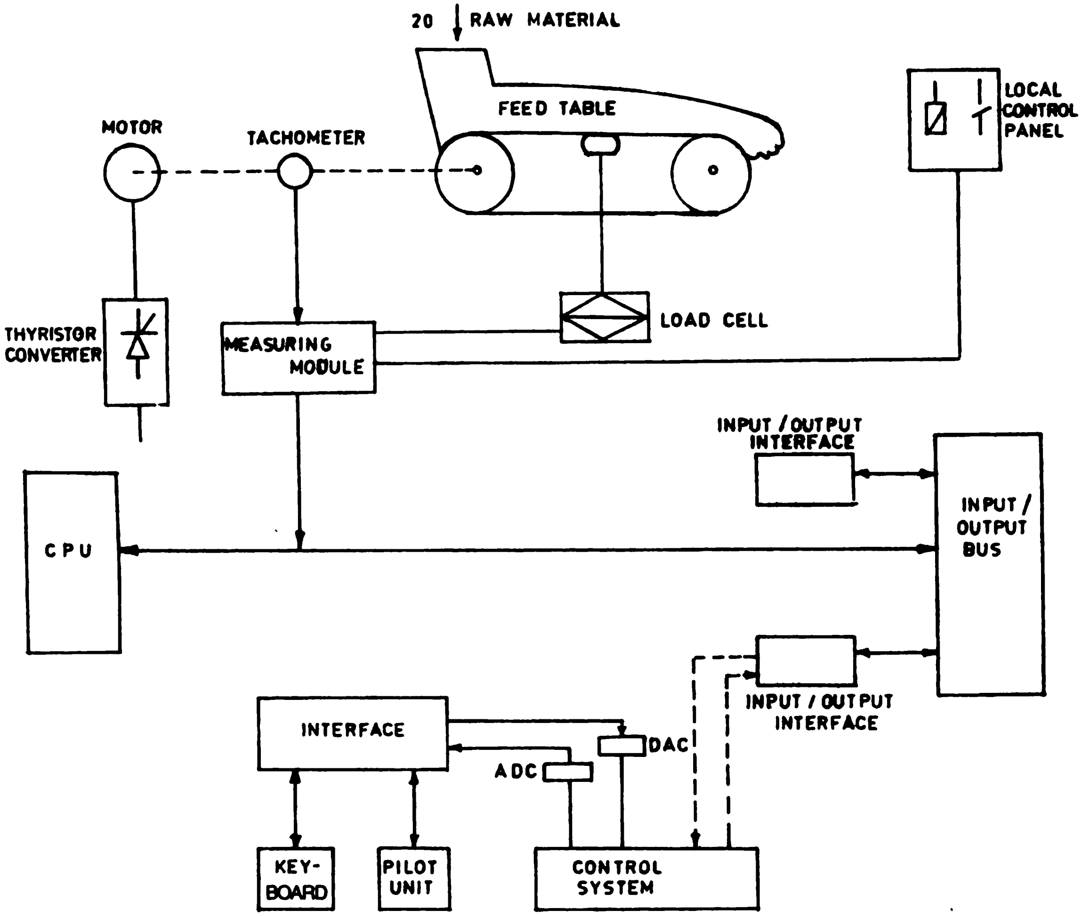


Fig. 8.6 Schematic diagram of a weighfeeder.

Normally, gearboxes above 3 MW rating are not available. So, for large ratings of the mill exceeding 3 MW, twin drives are employed having identical motors and starters.

8.3.3 Separator

The output of the ball mill is transported to the separator in which fine and coarse cement are separated. A rotor is rotated in the separator. The speed is controlled between 150–300 rpm to adjust the fineness of the cement product. A converter-fed dc motor/variable speed slip-ring induction motor/variable frequency inverter-fed induction motor may be employed. The speed range is 1:10 by armature voltage control at constant torque. The drives are normally designed for six starts/stops per hour in cold condition and four starts/stops per hour under hot condition.

8.3.4 Blower Drive

Squirrel cage induction motors with start/stop push buttons, overload relay, fuses and contactors are used for all blower drives.

8.3.5 Cyclone

The gas output of the separator is carried to a device called the *cyclone* in which the cement product is separated from air dust. The cement product is transported to silo. The air dust always contains a certain amount of fine cement. So, an air dust is taken to an electrostatic precipitator from which clean air goes out through the chimney and fine cement particles are taken to the silo. The cyclone drive is normally a slip-ring induction motor.

8.4 ELECTRIC TRACTION

Thousands of commuters all over the world get the benefit of EMU (Electric Multiple Unit) services regularly. In suburban railways, EMU service is very useful because of its advantages of quick start and stop enabling a high speed schedule. Electric traction system is without a parallel for underground as well as main line service. The system is very efficient, clean and pollution-free. But the major disadvantage lies in its capital cost, which is very high for a metro railway system. Power failure even for a short period can lead to complete disruption of the running schedule of the trains.

8.4.1 Tractive Effort

Figure 8.7 shows the driving mechanism of an electric traction system. Tractive effort is applied at the rim of the driving wheel through gears to propel the train. The gear ratio and wheel diameter

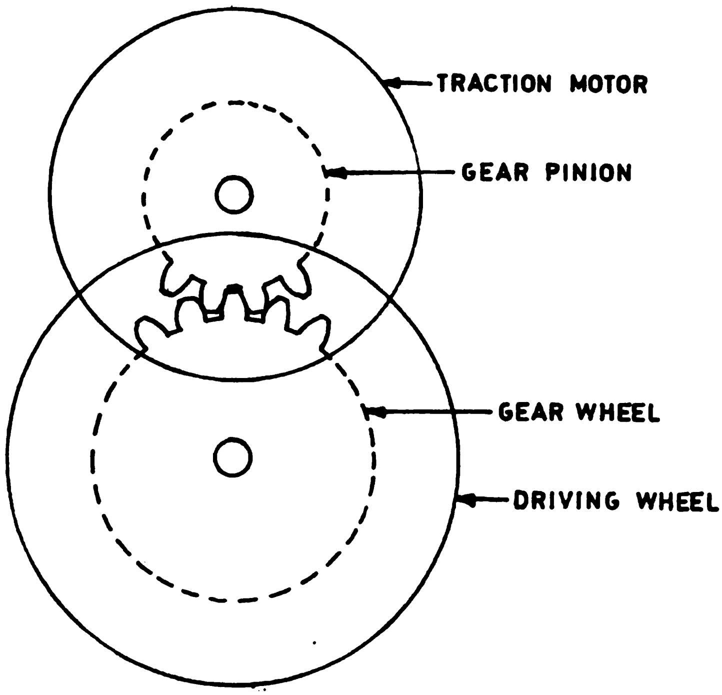


Fig. 8.7 Driving mechanism of an electric traction system.

are fixed. So, the tractive effort is proportional to motor torque. To provide a large tractive effort, more than one motor is normally used. The total tractive effort depends upon the number of traction motors.

8.4.2 Requirements of Electric Traction

(a) Traction equipment should be robust and sturdy enough to withstand continuous vibrations, dust and humid environment.

(b) Power to weight ratio of the traction motor should be high so that it occupies less space.

(c) High tractive effort at starting.

(d) It should be possible to overload the motor for a short period.

(e) Ability of traction motors to apply regenerative braking during descent.

(f) Coefficient of adhesion should be high.

(g) The traction motors must be capable of withstanding voltage fluctuations and interruptions of power supply.

(h) The motors should be amenable to simple speed control methods.

8.4.3 Suitability of Series Motors

All the above requirements of electric traction are suitably met by dc series motors. A dc series motor is, therefore, the ideal choice. The motor has to provide sufficient tractive effort (TE) to set the train in motion. Figure 8.8 shows the total resistance offered by the windage and friction to the motion

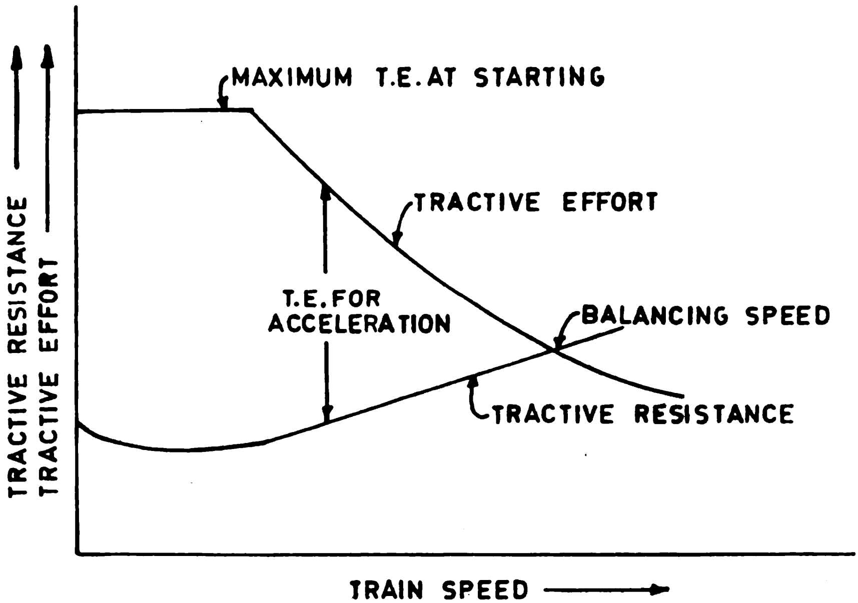


Fig. 8.8 Train acceleration.

of the train along with the series motor characteristic. The maximum TE is produced at starting when the motor is separately excited. The difference between the TE offered by the motor and the tractive resistance is the force available for accelerating the train. As the train gains speed, TE gradually decreases and eventually the balancing speed is reached. It is, therefore, seen that a series motor offers the best speed-torque characteristic for electric traction.

The heavy conductor sections used in the armature and field circuits make the series motors robust and reliable. Constant horsepower operation takes place. The motor current is restricted to a certain permissible limit. The motor can draw over-current for short periods. The motor efficiency through gears is normally high, above 80%. In traction systems, both electrical and mechanical forms of braking are applied. Electrical braking offers smoother and higher retardation than that obtained in mechanical braking. The only disadvantage with series motors for electric traction is that regenerative braking is not readily possible. A series motor has to be separately excited for regenerative braking. It may be noted that a dc series motor has a higher power to weight ratio than that of an ac series motor, which also gives commutation difficulties at low speeds.

8.4.4 Coefficient of Adhesion

It has been stated earlier that tractive effort (TE) is proportional to the torque exerted by the traction motor. But there is a maximum TE beyond which the above relation does not hold good. The maximum TE is proportional to the dead weight.

$$(\text{TE}) \propto W = \mu_a W$$

where μ_a is called the coefficient of adhesion, which should be high to produce high TE and acceleration of the train.

8.4.5 Supply Systems and Traction Motors

There are two types of electric traction systems: (a) ac traction, and (b) dc traction.

AC traction. AC series motors are used. Commutation difficulties are experienced on standard industrial frequency supply at low speeds because of circulating currents in the short-circuited coil. That is why ac traction motors are operated at a lower frequency, e.g. $16\frac{2}{3}$ or 25 Hz, the rated voltage being 415 V. So, a transformer is required, if the supply is obtained at low frequency dedicated to the traction purpose; otherwise, a frequency converter is required to be carried on the train.

DC traction. DC series motors are used. The advantages of using dc series motors have already been stated. In this system, the dc operating voltage is 1500 V for urban and suburban services and 3000 V for main line service. The dc voltage is restricted to 3000 V to avoid commutation problems. Thyristor converters and choppers are used for driving dc traction motors resulting in high efficiency, reliability and smooth control. A transformer with tappings and on-load tap changers are also carried on the train to step down the voltage.

In both types of traction systems, electricity is collected from the overhead contact wire at 25 kV with the help of a pantograph. Adequate insulation has to be provided to protect the train from electrocution. The rail line serves as the return path for current.

Traction substations are set up enroute from one terminus to another terminus station. Assuming that 132 kV source is available, the substations house a 132/25 kV step-down transformer with proper switchgear and protection systems. In case of suburban lines, the substations are spaced 3 to 5 km apart, whereas for main line they are spaced 40 to 50 km apart. For low frequency operation, the substations may be spaced at a distance of 50 to 80 km apart. High voltage is preferred for overhead contact wire line so that its conductor section is reduced and the spacing between conductor sections increased. The ac system is preferable for overhead contact wire system from the point of view of current distribution. In metro railway systems, dc supply is given between the running rails.

8.4.6 Train Movement

A train needs to be run optimally between the stops from the point of view of energy consumption, considering the weight carried and the distance traversed by the train. Speed-time and speed-distance curves are needed for the study of optimal train movement. In general, while moving from one station to another, the motion of the train may be of five types as shown in Fig. 8.9.

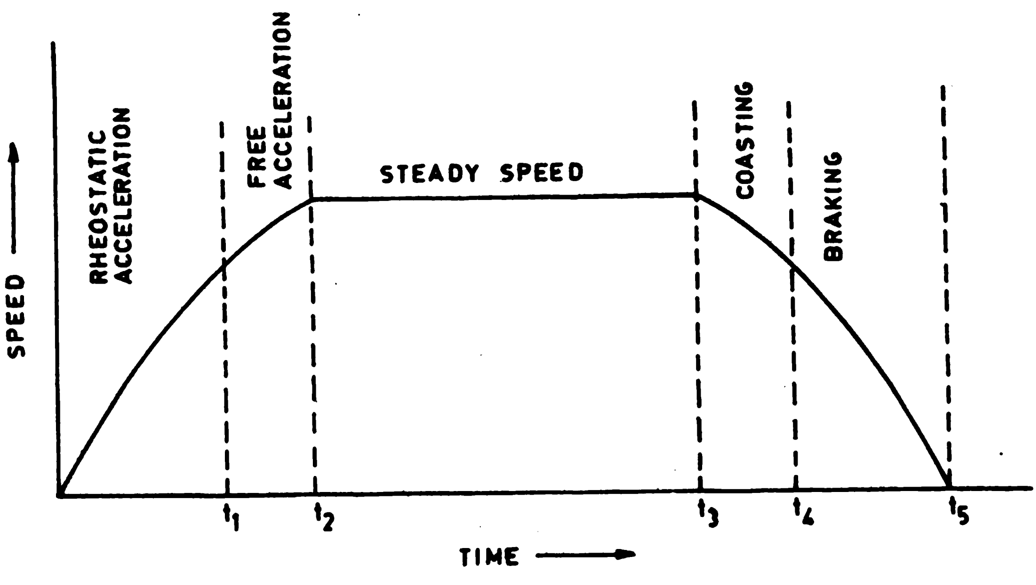


Fig. 8.9 Classifications of train movements in electric traction.

(1) During rheostatic acceleration, the starting resistance is cut out step-by-step so as to maintain constant current and constant acceleration.

(2) At the instant t_1, all starting resistance has been cut out. The train continues to accelerate up to the instant t_2, when balancing or steady speed is reached.

(3) The train runs freely at steady speed in this section.

(4) When the train approaches a stop, the power supply to the traction motors is switched off The speed falls, but the train continues to run due to momentum. This duration is called the *coasting period*.

(5) First the electric brake and then the mechanical brake is applied to stop the train.

8.4.7 Layout of Electric Drives

The power for electric traction is provided by traction motors to the driving wheel through gears. The traction equipment, required for driving and controlling the train, is mounted in the underfloor bogies. A suburban or metro train consists of a number of units. Each unit consists of a motor coach, a trailer coach and a driven trailer coach. The traction motors are mounted under the trailer coach. The whole train can be operated by one driver from either end. Electric traction is more suitable for suburban trains where frequent start, stop and high speed schedules are required.

Electric locomotives are used for long distance trains. An underframe superstructure is mounted in two bogies of the electric locomotive. The traction motors and their control equipment are mounted on the superstructure. The roof of the locomotive is made strong to hold the pantograph. The driver's cab is also built in the locomotive with all controlling gadgets. The driver's control is for hand regulation. There is a master controller in the driver's cab to regulate train operations, e.g. starting, acceleration, speed control, coasting, braking, reversal of motors, etc. As the traction motors are mounted under the floor of the motor coach or the locomotive, they should be suitably protected from dust and humidity.

In a diesel electric locomotive, a motor-generator set is also carried in the locomotive to provide power to traction motors, which drive the wheels of the locomotive. Series generators and series motors are used. Diesel locomotives are suitable for hauling goods trains as they provide a high tractive effort to pull the large number of wagons involved. Two electric locomotives are sometimes deployed to run a goods train.

8.4.8. Speed Control

The speed of traction motors is controlled by the combined series-parallel method, and rheostatic method. Section 3.3.1 may be referred to in this connection.

In electric traction, four or six identical motors are used. Constant horsepower operation takes place. There are separate notches of the master controller for different speeds. For long distance trains, acceleration and braking periods are negligible compared to the running time. On the other hand, free running is almost absent for suburban trains. There may be different arrangements of motor connections yielding different speeds. For example, a six motor locomotive having each motor rated for half the line voltage may undertake the following connections (A set comprises two motors connected in series):

(a) Three sets connected in parallel yielding full speed.

(b) Three sets connected in series yielding one-third full speed.

If three motors are connected in series to form two sets across the line voltage, two-third full speed is obtained. Similar interconnections may be made with four motors to have different speeds.

Accelerating resistance is used for adjustable voltage control of dc traction motors placed under the carriage. For getting various speeds, transformer output voltages are also adjusted with the help of on-load tap changers.

8.5 COAL MINES

8.5.1 Precautions

The environmental features in the underground of a coal mine present the most hazardous conditions of work to miners as well as machines. It is a dark, dusty and damp environment with waterlogging at numerous places. There is no natural circulation of air. The most dangerous of all is the fire hazard, which may be caused by any form of sparking, flash or flame, if the inflammable gas or vapour (methane, CH_4) produced at the coal face exceeds 0.5% in the ventilating air. Carbon monoxide (CO) is the killer gas, which is produced due to insufficient burning of methane. The working conditions in the underground are thus dangerous. An outline of the precautionary measures adopted in respect of electrical systems in the mines to combat the killer gases is given below.

(a) Maximum limits of working voltages in a mine have been prescribed by Indian Electricity Rules (vide Rule no. 118) as under:

1. Electrical energy shall not be transmitted to a mine at voltage exceeding 11,000 V and shall not be used at a voltage exceeding 6600 V.

2. The voltage is stepped down to 6600 V, 3300 V, 1100 V, 550 V, 230 V, 110 V and 30 V according to the requirement for the equipment of the mine. It may be noted that the working voltage in the underground of mines is 550 V and not 415 V.

3. The voltage shall not exceed 1100 V for portable and transportable electrical machines. Normally, the motors are rated at 550 V for hp ratings up to 100 hp, and are rated at 1100 V for hp ratings between 100 to 500 hp. On the surface of the mine, the motors may be rated at 3.3 kV for 400 hp and above. Motors rated above 500 hp, are rarely used in the underground of a mine. The choice of the rated voltages for drives on the surface of the mine is dictated by the supply available, the size of motors and economic factors.

4. For lighting systems, the voltage shall not exceed 125 V, normally 110 V.

5. For control circuits, the working voltage is 30 V.

(b) All electrical equipment shall be **flameproof (IS 2148)** for safe operation in the underground. where inflammable gases may exist. The flame and spark created inside the enclosure shall not be allowed to ignite the gases outside the enclosure. The created pressure inside the enclosure shall not be able to destroy the enclosure, which should be of sufficient mechanical strength. Totally enclosed fan-cooled or totally enclosed water-cooled (TEFC or TEWC) motors are recommended for use in mines. The commonest form of protection is flange protection, whereby gases or flame produced in the motor come in contact with the metal of the flanges and emerge at a temperature below the ignition temperature of the inflammable gas or air outside.

A circuit in which any sparking or heating is incapable of causing ignition of a prescribed gas or vapour is called the *intrinsically safe circuit (IS 5780–1980).*

(c) Methods of neutral grounding used in mines are as under:

1. **Solid grounding.** This system is called the Earthed Neutral System, which is used in almost all conventional underground mines in India.

2. **Impedance grounding.** An impedance is inserted between the neutral and the ground. This system is called the Restricted Neutral System.

3. **Ungrounded system.** This system known as Isolated or Insulated Neutral System is used in countries like Germany, France and Russia. It is a continuous monitoring insulation system in which risks of fire and human shock are very low.

(d) As per Coal Mines Regulation (CMR), there should not be any electrical power in the underground until the ventilation fan is on.

(e) Liquid resistance is used in the underground to eliminate fire hazard. The electrolyte is a solution of sodium carbonate in distilled water, whose composition varies from 3 to 10%, i.e. about 50 gm of salt in one litre of water.

(f) Five-core flexible cable is used for coal-face machineries, e.g. coal drill, coal cutting machine, coal dumper, etc. For all other motors, three-phase double armoured mining type cables are used. The example of a drill motor is cited in Fig. 8.10. When the drill switch is closed, the pilot relay is energized, which in turn closes the line contactors giving supply to the drill motor. It facilitates remote control. This is particularly advantageous as the operator handles very low voltage.

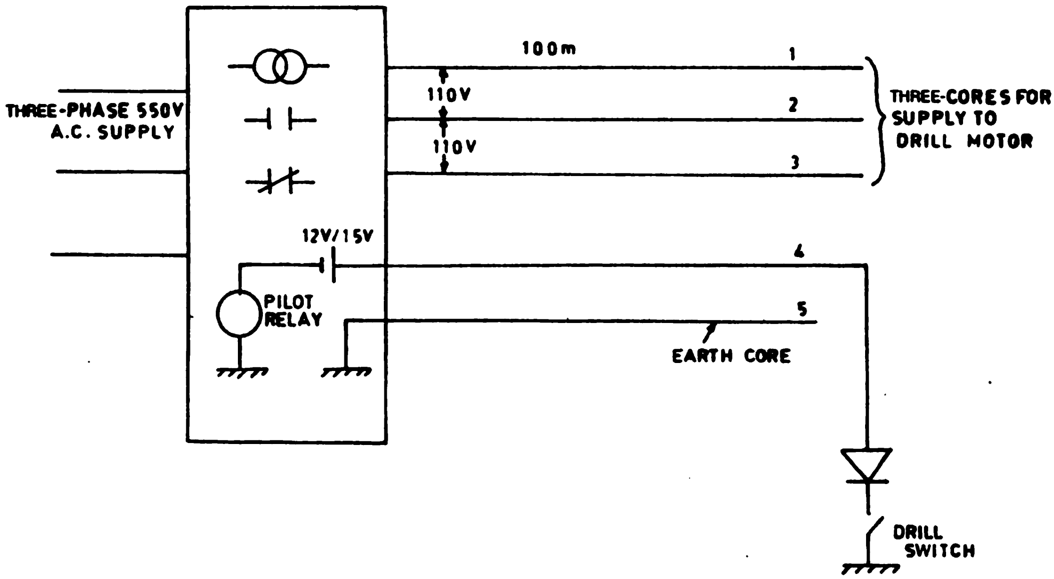


Fig. 8.10 Five-core flexible cable connection for a drill motor used in mines.

(g) The insulation of the motor shall be minimum of class F. The winding and insulation must be capable of withstanding hot and humid conditions for a long trouble-free life.

8.5.2 Drum Winder

The access to coal face in a deep mine is by a well called the pit or shaft (Fig. 8.11). Access by incline is cheaper and quicker than by shaft. Incline passage is suitable for shallow mining operations. An incline mine cannot be economically justified for deep mining operations because a long slope length is required.

Several types of winding systems are sketched in Fig. 8.12. The drum operates two conveyances of which one is raised, while the other is lowered. Audible or visible signals can be transmitted between the bottom, banking level and the winder house. The winder can be started only when the shaft is free. The position of the cage in the shaft is monitored by Selsyn transmitters and receivers. A phase sensitive detector finds out, whether the cage is going up or down. All types of winders are manually controlled as per coal mine regulations. Fully automatic control of winders is permitted

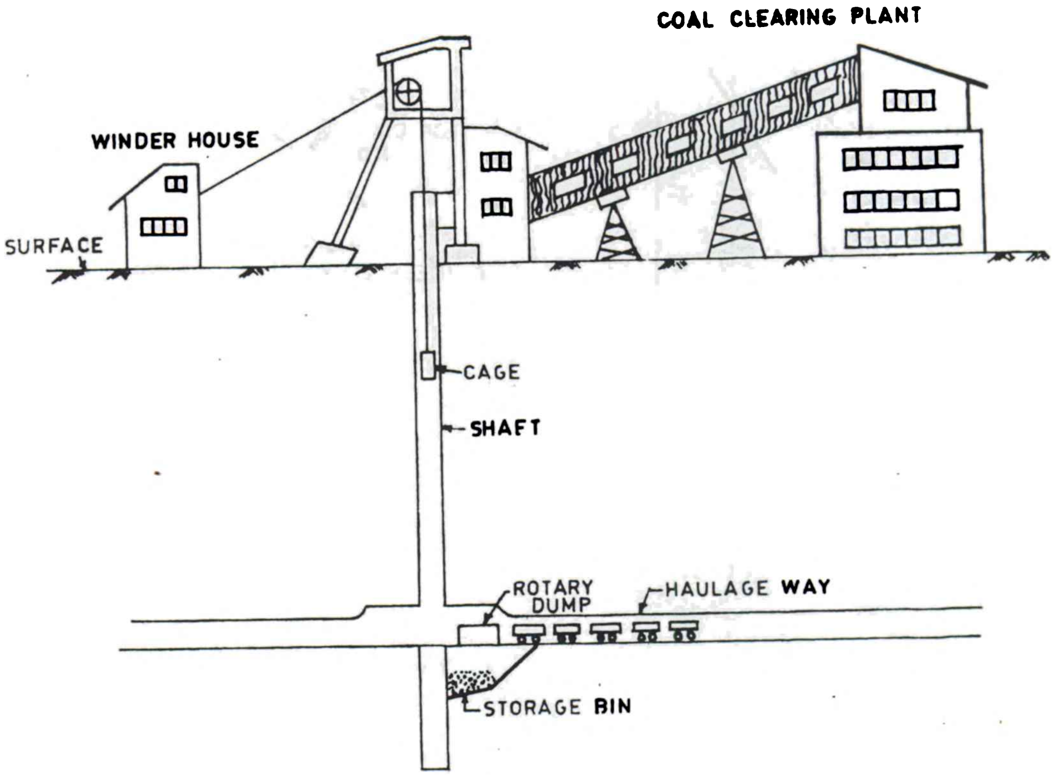


Fig. 8.11 Shaft in a coal mine.

only for skip winders with constant load for transport of materials only. The requirement for control of a winder is estimated from its duty cycle. A typical duty cycle of a mine winder is shown in Fig. 8.13.

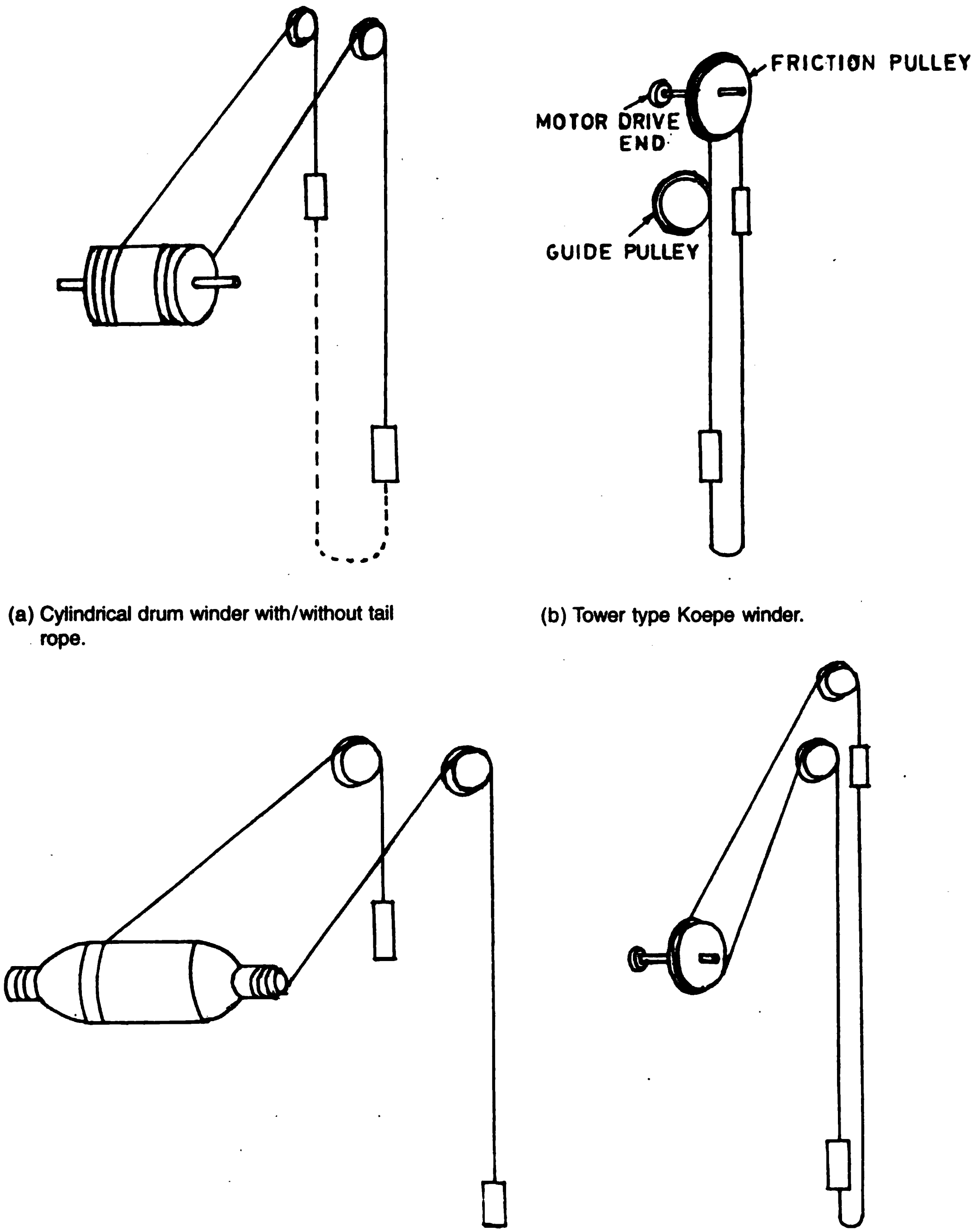
8.5.3 Modern Winder Drives

Three types of drives are employed for control of winders:

(a) Slip-ring induction motor drive
(b) Ward-Leonard dc motor drive
(c) Converter-fed dc motor drive

Slip-ring induction motor drive

Slip-ring induction motors are used for small winders upto 500 kW rating. The common method for speed control comprises insertion of liquid resistance into the rotor circuit. Single-phasing prevention, over-current relay, earth fault relay and arc blocking are provided in addition to the circuit breaker and forward and reverse contactors. For a small motor of about 200 hp, drum controller and metallic grid rheostat may be used. All induction motor drives incorporate reduction



(a) Cylindrical drum winder with/without tail rope.

(b) Tower type Koepe winder.

(c) Bi-cylindroconical drum winder.

(d) Ground type Koepe winder.

Fig. 8.12 Types of mine winders.

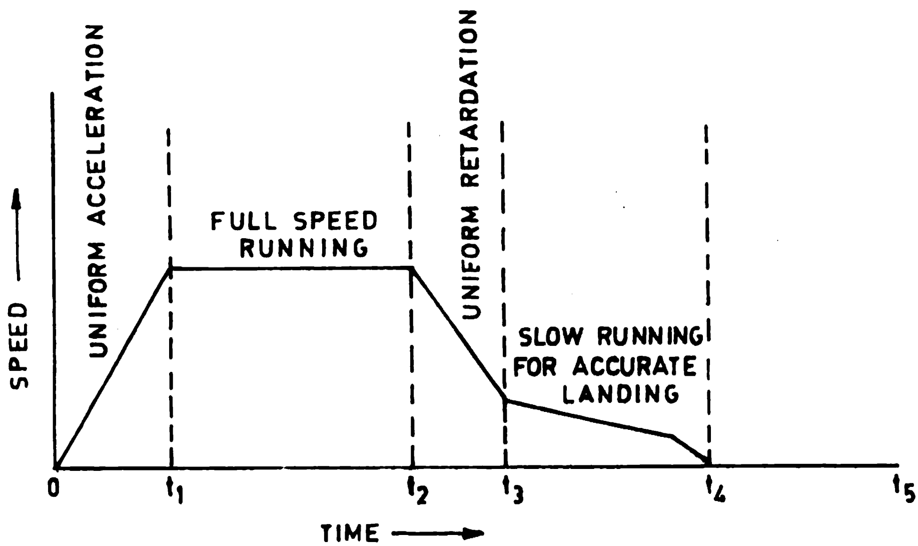


Fig. 8.13 Typical duty cycle of a mine winder.

gears, single or double. The retardation control is achieved by dynamic braking, i.e. by dc injection. Alternatively, a cage induction motor in association with a cycloconverter may be employed for control of the winder.

For widely varying torque demand and accurate decking, the closed-loop control is favoured. The rotor resistance is automatically controlled by a servomotor, which acts on driver's lever-controlled potentiometer speed reference, tachogenerator feedback and current feedback.

Ward-Leonard dc motor drive

DC winders are employed for higher capacities on account of the following advantages:

(a) Smooth and perfect speed control in either direction is obtained by Ward-Leonard method.

(b) Synchronous motor-generator set serves the purpose of power factor improvement.

(c) The maximum kVA demand and cost per wind are both less for a dc winder than for an equivalent capacity ac winder for deep shafts.

(d) Gear drive is not required.

As regards closed-loop control, the dc winder associates an amplidyne for three-loop control for current feedback, speed feedback and position control.

Converter-fed dc motor drive

In this method, the motor-generator set of Ward-Leonard control is replaced by a three-phase full-controlled thyristor bridge converter. It operates at lagging power factor. The amplidyne is also replaced by an electronic amplifier for feedback control. Electronic operations ensure safety, reliability and higher productivity. Microprocessor-based control is used in modern winding systems. Permissible speed/depth characteristic is stored in the memory. It facilitates better precision in optimal control.

8.5.4 Shearer

In long wall mining, the drum shearer is employed to cut out and load the coal into an armoured face conveyor (AFC), which traverses along the coal face in concert with coal cutting. The shearer assembly is mounted on a robust underframe on the armoured face conveyor. There are two cutting drums at each end of the machine. These drums are carried on the ranging arms fitted to the handed gearheads. The shearer automatically traverses as the cutting progresses. The armoured face conveyor along with the shearer can be advanced without dismantling. The shearer haulage (traction) speed is automatically adjusted depending upon the load on the main driving motor. There may be single or double motor arrangements to drive the shearer. For obtaining high output in long wall operations, the standard motor configuration includes double cage induction motors (2 × 375 kW). Most of the power is consumed for coal cutting, whereas a portion of it is used for its haulage/traction. The coal face is sprayed with water by a separate pump motor for suppression of coal dust. Single or two-speed gearhead boxes provide a wide range of speed control of the cutting drums. The shearer motor is robustly constructed to withstand frequent starting, which is a common feature for any mining motor. The rotor of the motor is normally air cooled by fitting a fan on its shaft and the stator is water cooled. The stator frame is constructed as a double-walled cylinder for placing a water jacket around it. Thermostats are used to trip the power supply, if the temperature of water exceeds the maximum specified limit. Normally the shearer motor cannot be restarted without the intervention of the operator, if the power supply goes off.

Drum shearers of two makes are known to be employed in coal mines: (a) UK make, and (b) Russian make. These two makes mainly differ in respect of their haulage systems; besides, the UK make shearer has one driving motor, while the Russian make has two. Otherwise, the general description of the shearers is the same as stated above.

UK make shearer

The power to traverse the shearer along the coal face is provided by a servo-controlled, closed circuit hydrostatic system (Fig. 8.14) driving a sprocket, which engages in the links of the haulage chain. A single motor rated at 1100 V, 375 kW is used. Infinitely variable speed is obtained. Haul/left/right/stop operations are obtained by the rotation of the control handle.

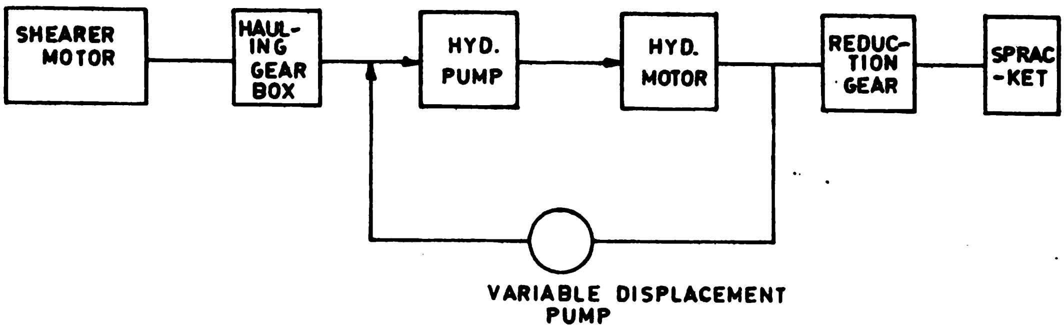


Fig. 8.14 Haulage system of the UK make shearer.

Russian make shearer

The block diagram of the shearer drive is shown in Fig. 8.15. When the depth of cutting is high,

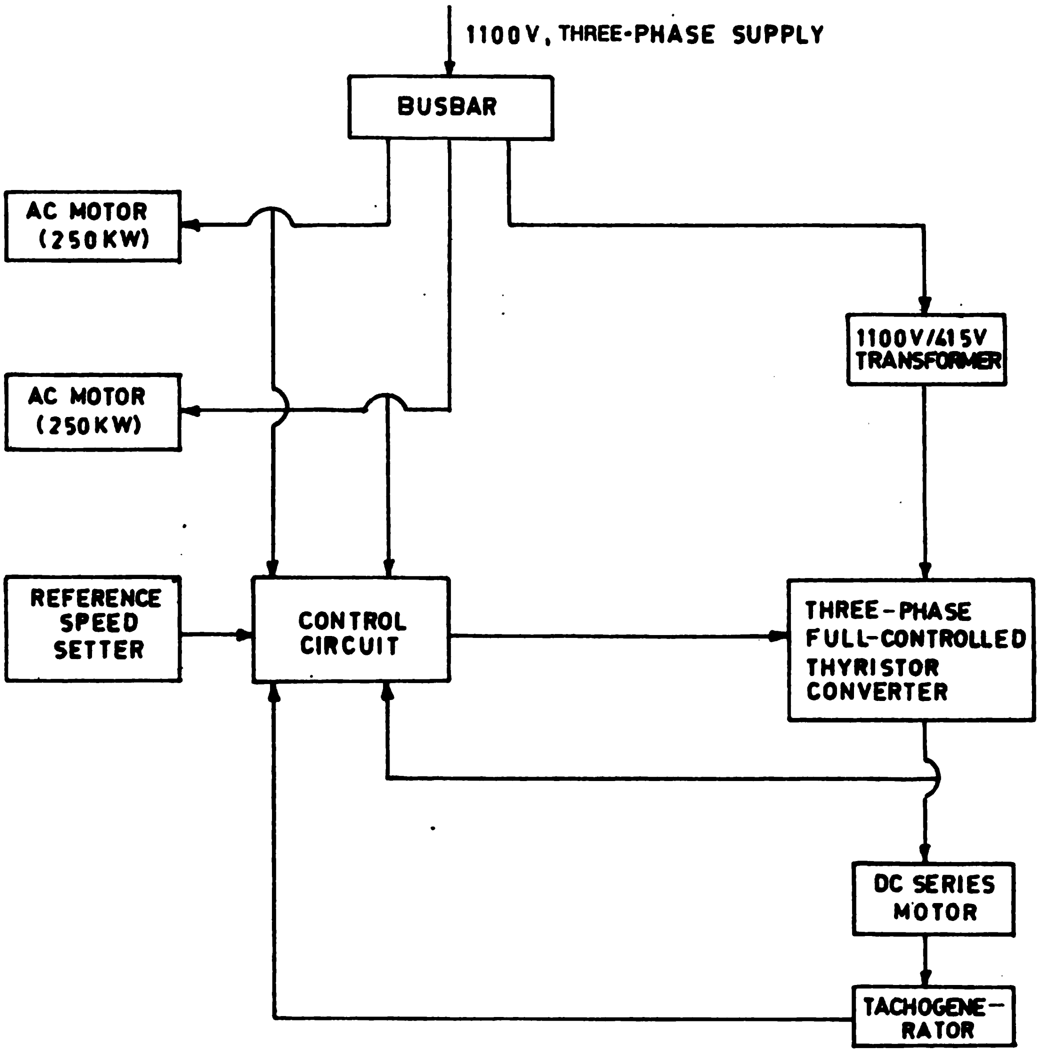


Fig. 8.15 Schematic diagram of the Russian make shearer.

the traction speed is automatically lowered by current feedback. Two double cage induction motors are used in this system. The cooling system of these motors is the same as in the UK make shearer.

8.5.5 Conveyors in Long-wall Systems

The coal is transported from the coal face to the bunker by (a) an armoured face conveyor, (b) a bridge stage loader, (c) a gate belt conveyor, and (d) a main belt conveyor, placed in the same order.

550 V, 112 kW, four-pole, flange mounted, continuous rated, single speed, flameproof, fan cooled, class F insulated, DOL started, three-phase cage induction motor with minimum starting torque and minimum pull-out torque equal to 175% and 170% respectively, may serve as standard

specification for the above mentioned conveyors. The voltage and power ratings may vary depending upon the load to be transported.

8.5.6 Auxiliary Motors

Auxiliary motors may have the following specifications: 7.5 kW/4 kW, 550 V, three-phase 50 Hz, four-pole, Class F insulated, flange mounted, DOL started, continuous rated cage induction motor to IS 2148.

8.6 PAPER MILL

8.6.1 Pulp Manufacture

Paper is made from pulp, which is a soft fibrous mass. Wood, bamboo and jute are the chief raw materials. The following operations are undertaken for making pulp from the raw materials.

1. The raw materials are cut to pieces with a chopper for which the slip-ring induction motor may be employed without any specific control.
2. The chips are crushed in a grinder. The capacity of the grinder ranges from 3 to 4 MW. A slip-ring induction motor/auto-synchronous motor with gear drive is used.
3. The crushed materials are cooked and digested by chemical treatment. The cellulose pulp fibres are produced.
4. The bleaching operation is undertaken for washing, cleaning and whitening of the fibres.
5. Pulp is then further processed by beating and refining before formation into paper. There is no sharp difference between these two operations. The fibres are cut to tiny pieces in this process to improve the quality of the paper. For this operation, a slip-ring induction motor or an auto-synchronous motor is employed. No sophisticated control is required.

8.6.2 Paper Manufacture

Paper is a web of cellulose fibres. The fibres are uniformly suspended in water with consistency (solid content) 0.5–1%. The paper is made from these watery pulp fibres. The production line in a paper mill comprises the following five sections, where web is the physical link between the sections.

(a) Wire or Couch section

(b) Press section

(c) Dryer section

(d) Calender section

(e) Reel section

Wire section

The watery pulp fibres are transferred to a fine wire mesh, which moves on rolls. In this section of the paper mill, the fibres are uniformly deposited on the wire mesh while the water drips through it. The formation of a sheet takes place in this way. The removal of water from the sheet is enhanced by the action of suction boxes placed under the wire mesh. The schematic diagram is shown in Fig. 8.16.

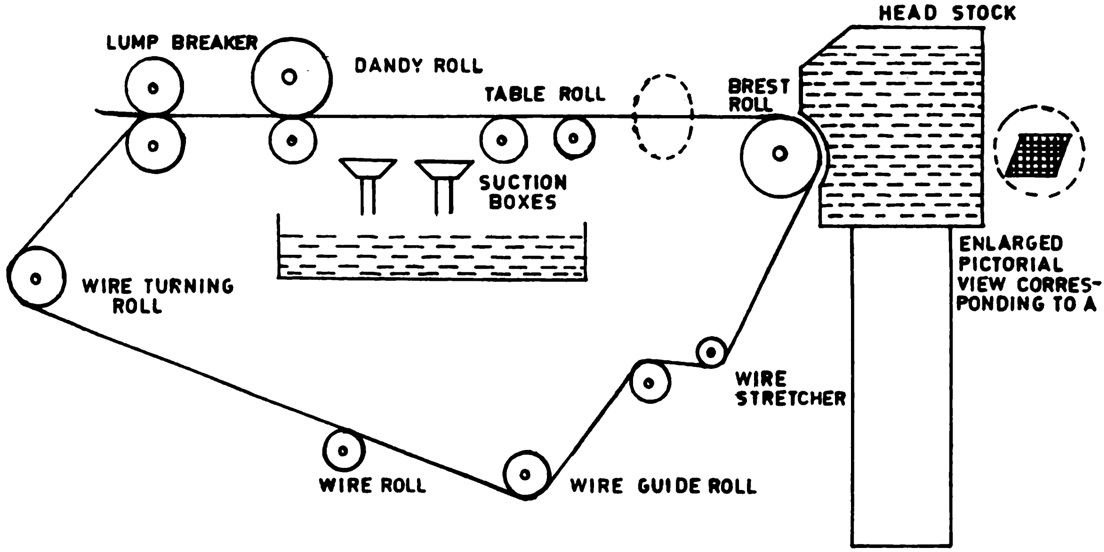


Fig. 8.16 Schematic view of the wire section in a paper mill.

There are two types of modern wires, which differ basically in their de-watering system: fourdrinier wire (Fig. 8.16), and twin wire. The former requires more energy than the latter because of the action of the suction boxes. De-watering of the twin wire takes place by the action of centrifugal force in one or two bends formed by suction rolls. The combination of the above two methods of de-watering is also possible. The rolls constitute a form of conveyor system. The dandy roll is used for water-mark. The wire-turning roll is speed controlled. The other rolls, such as breast roll, wire-guide roll, are non-controlled. These are helper drives to the main wire-turning roll drive. The suction rolls are current controlled.

Press section

In the press section (Fig. 8.17), the web from the wire section is pressed to squeeze the water out from the wet sheet. At the output stage of the press section, the moisture content is reduced to 60–65%. The maximum power is required in the press section of the paper mill. One of the press rolls is speed-controlled, while others are torque-controlled. Felt-guide rolls, pick-up rolls, etc. are non-controlled helper drives.

Dryer section

The press section is followed by the dryer section, where the web passes over and under the heated cylinders. The paper is dried up in this process until the moisture content is reduced to about 6%. The tearing of paper mostly takes place in this section because of uneven drying of the paper web.

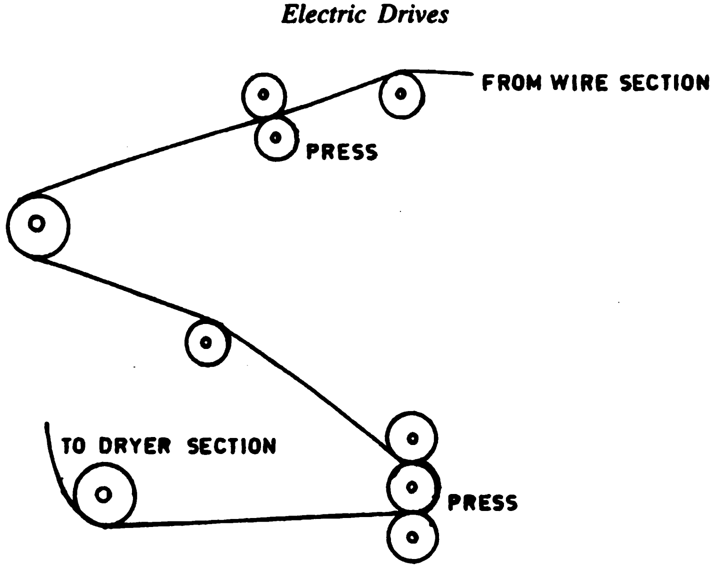


Fig. 8.17 Press section of a paper mill.

The variation of tension also occurs here. Therefore, the control of the electric drives in the dryer section is of utmost importance. A sketch of a dryer section is shown in Fig. 8.18. Each roll has its own drive motor, which is speed as well as current controlled. A converter-fed dc shunt motor is used.

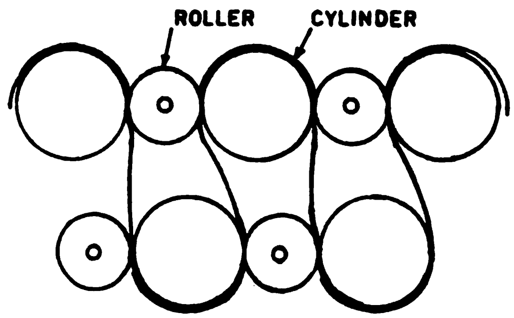


Fig. 8.18 Dryer section of a paper mill.

Calender section

In the calender section, the paper is subjected to pressure and friction so as to produce a smooth and glossy surface. The super calender (Fig. 8.19) is a three-section machine, e.g. reel, calender stack and unwind. The last dryer section may also be located immediately before the calender stack. Each of these sections has a drive motor. The calender stack operates as the master section having speed control with tachogenerator feedback. For reel or centre-wind drive, the speed is controlled by a voltage regulator having an inner current loop with web tension control. The web tension feedback may be obtained with the help of load cells.

Pope drive

Pope drive (Fig. 8.20) is a peripheral, which is used to reel up the finished paper. The drive motor

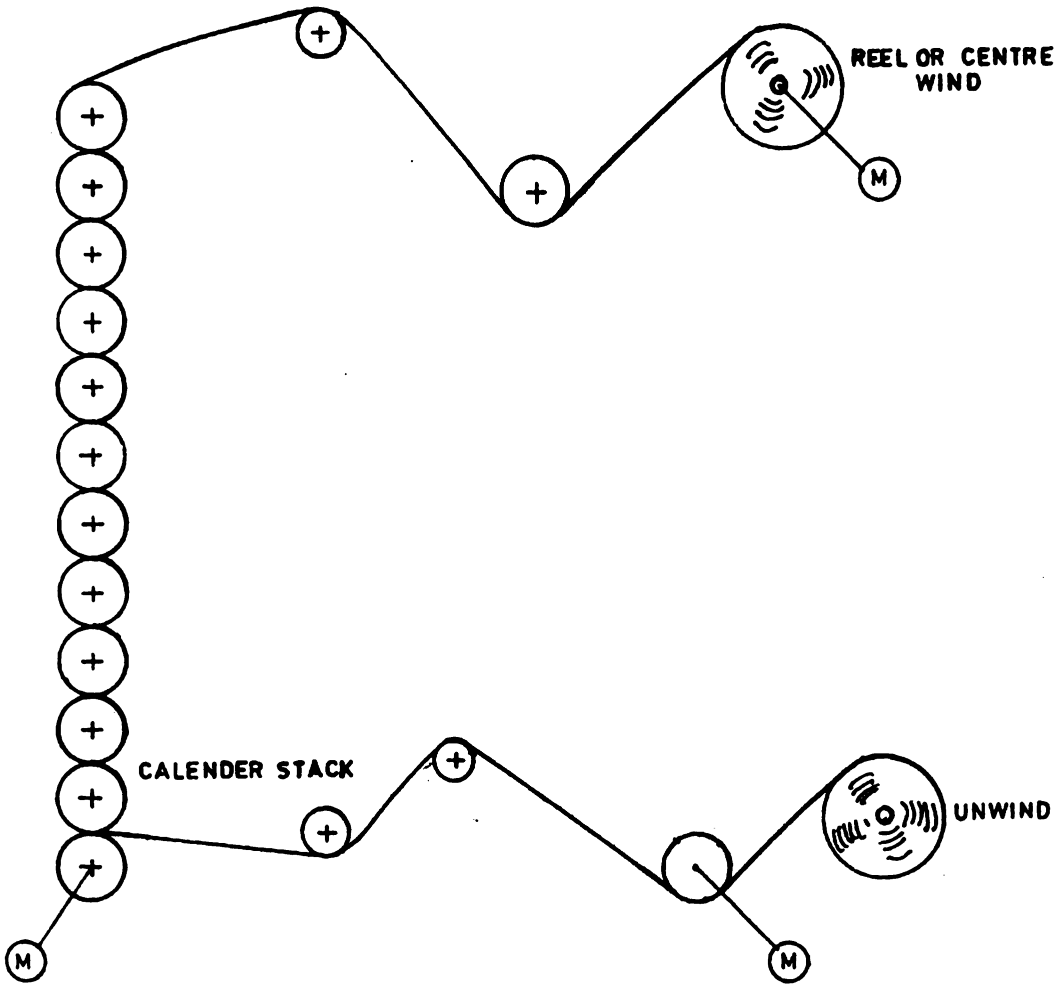


Fig. 8.19 Calender section of a paper mill.

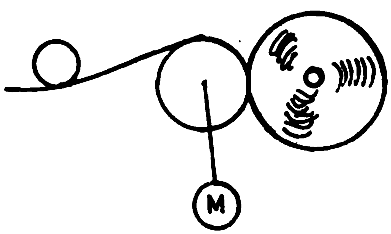


Fig. 8.20 Pope drive.

is speed as well as current controlled. The speed control comes into action only, when slipping of pope roll occurs. Otherwise, the control is the same as for helper drives in wire and press sections.

8.6.3 Paper Mill Drives and Control

It is hard to maintain close coordination between the various sectional drives as the material is weak. The weight of the paper as well as its quality is governed by the speed of the paper machine. The motors used in the paper mill are splash water protected, pipe ventilated, dc separately excited (shunt) type with class F insulation. An electronic controller with high amplification factor and fast response is called for.

Each section of the paper mill has its own speed regulator. Three-phase full controlled thyristor bridge converters are used. The controller comprises an outer speed regulating loop and an inner current regulating loop. The current feedback is taken from the current transformer on the input side of the converter.

The control of a paper machine can be accomplished in two ways, e.g. analog and digital. The reference speed is set by a potentiometer. There are provisions for start, stop, crawl and jogging operations in the control circuit, which also includes soft start. Reverse inching operations of the sectional drives can be brought about by field reversal of the driving motors.

If there are a number of motors in a section, one of them operates as a lead motor, whose speed is controlled by variation of the firing angle of the thyristor bridge converter. The other motors are the helper motors, which are connected in parallel with the lead motor. The torque of the helper motor is adjusted by controlling the field current. A master controller is used for coordination of the speeds of various sections.

Digital control is preferred nowadays, because of its certain advantages over analog control. A digital control system gives better speed resolution and its superiority is such, that torque control and web tension control can be dispensed with. The quality of paper is closely monitored at the end of the calender section using a processor-based system with the help of which scanning is done crosswise over the moving paper. The information about speed, thickness, moisture content, etc. is collected, analysed and shown on a video display unit (VDU) of the system for data logging and visual displaying for operator (manual) intervention, if required, and also for providing the feedback needed for automatic control.

For sectional ac drives, splash water protected squirrel-cage induction motors with class F insulation are used. The PWM inverter-fed ac drive system with processor-based controller gives the widest control of speed and the best performance of the motor.

8.7 MACHINE TOOL DRIVES

The most widely used general-purpose metal cutting machine tools are lathe, drilling, shaping, milling, planing and grinding machines. Each machine has a different purpose and also uses a different machining method from the other. Various types of such machines are available. Apart from general-purpose machine tools, there are high production and special-purpose machine tools. Three-phase induction motors are used for driving machine tools in most cases. In a few cases, dc motors are used. Gears and stepped pulleys are used to provide a wide range of speeds. Pole-changing motors are used to augment speed variation. Two types of working motion are required in machining: (a) the primary cutting motion, which determines the cutting speed, and (b) the feed motion.

8.7.1 Lathe

A general-purpose lathe is driven by a motor, which is connected to the first shaft by means of a belt. Here, the work-piece, that rotates, is held in the chuck attached to the spindle. The power is transmitted to the spindle through a speed gearbox. The spindle speed must be variable as different materials of various diameters are cut. Different speeds are obtained by engaging different pairs of gears between shafts in the gearbox by means of clutches, sliding gears, etc. From the spindle, a branch of power transmission is taken through a gear train to the feed gearbox for rotation of a lead screw and a feed rod, which can move the carriage on which the tool is mounted. The feed gearbox is also provided with speed changing arrangements.

A representative diagram of a lathe is shown in Fig. 8.21. Further details, such as reversing gears, separate lead screw or feed rod, etc. are not shown. This scheme (Fig. 8.21) gives six spindle speeds, but in a general-purpose lathe more number of speeds can be obtained. Three-phase, 415 V, star-connected, 1 hp, DOL-started cage induction motor is used. Figure 8.22 shows a simple scheme for carriage movement on the lathe bed.

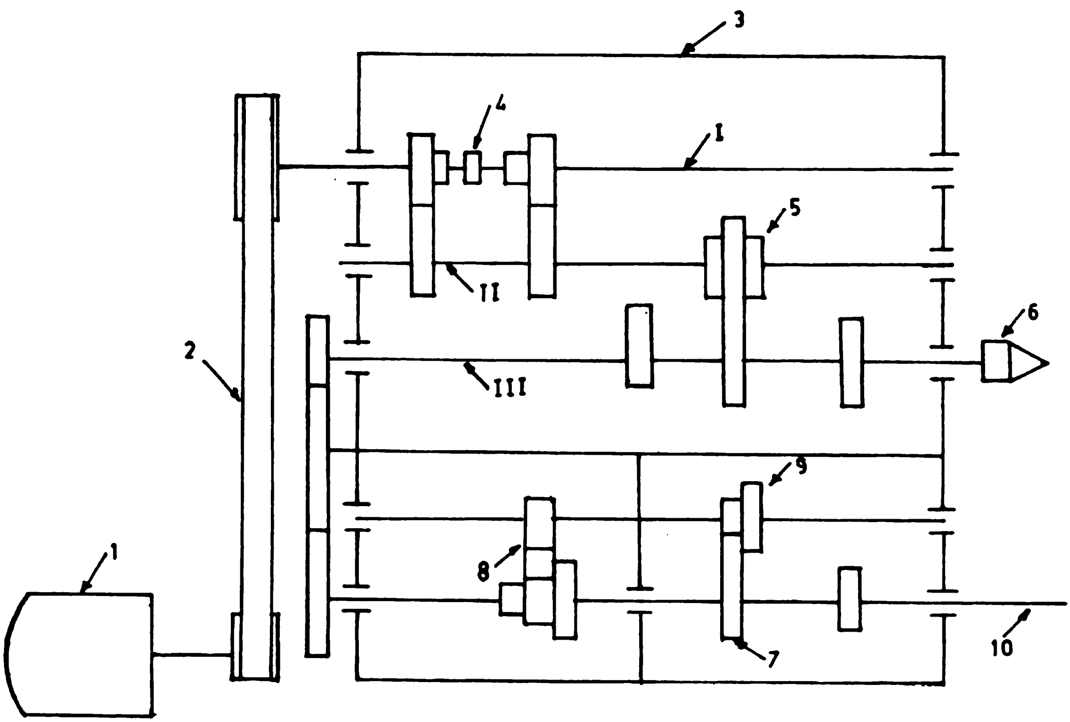


Fig. 8.21 Representative diagram of a lathe.

In order to have more number of speeds or to have the required number of speeds with a reduced gearbox, pole-changing motors are sometimes used. However, the speeds available are of step variation. These steps are in a geometrical progression. If one wants to obtain any speed between the maximum limits, a stepless speed drive can be used with the gearbox.

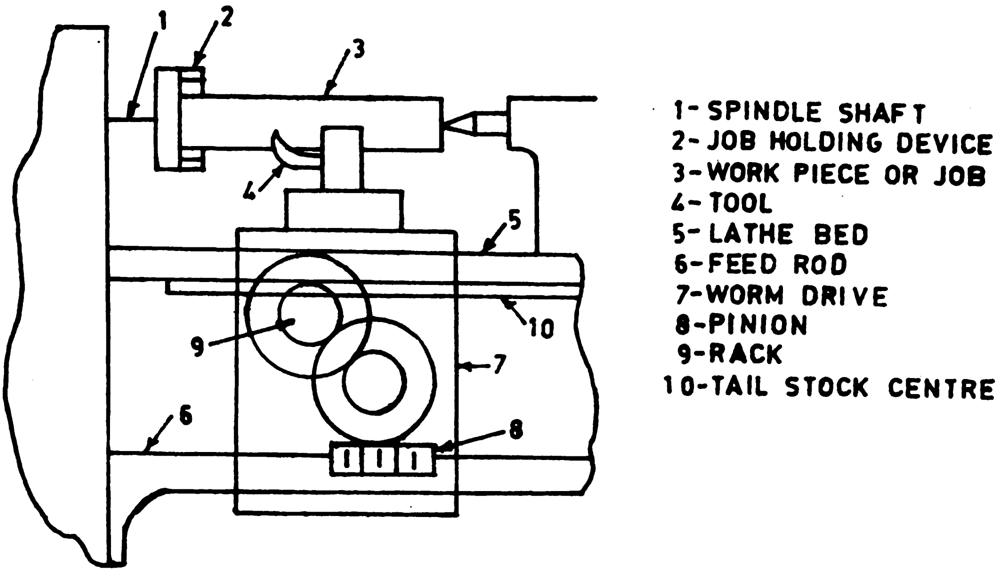


Fig. 8.22 Scheme for carriage movement on the lathe bed.

The following operations can be performed with a lathe: (a) Parallel turning, (b) Taper turning, (c) Facing, (d) Parting off, (e) Drilling, (f) Boring, (g) Threading, and (h) Knurling.

8.7.2 Drilling Machine

In a drilling machine, the spindle carrying the drill is vertical. Both speed and feed are powered from a single motor. In a radial drilling machine, the spindle carrying the drill head is mounted on a radial arm, which can swing about a column. For raising or lowering the arm, a separate motor is used. The motors used are three-phase induction motors with DOL starters. The speed control is obtained by gears and stepped pulleys.

8.7.3 Milling Machine

Milling machines are of various types: (a) Horizontal, (b) Vertical, and (c) Universal. Figure 8.23 shows an ordinary horizontal milling machine, where separate three-phase squirrel cage induction motors (SCIM) are used for spindle drive and feed motions. DOL or star/delta starters may be used. The speed control is achieved by gears and stepped pulleys.

The cutter is mounted on an arbor spindle and the job is clamped on the table. The speed gearbox is inside the column and the feed gearbox is within the knee of the body. The knee can be moved vertically along the guideways of the column. The table including the movement of the knee can be given three movements: longitudinal, cross and vertical. There are also handwheels and a crank for hand control of these feed motions. The vertical milling machine is another type having a vertical arbor instead of the horizontal one. In the universal machine, the table can be rotated so that it travels at an angle to the column.

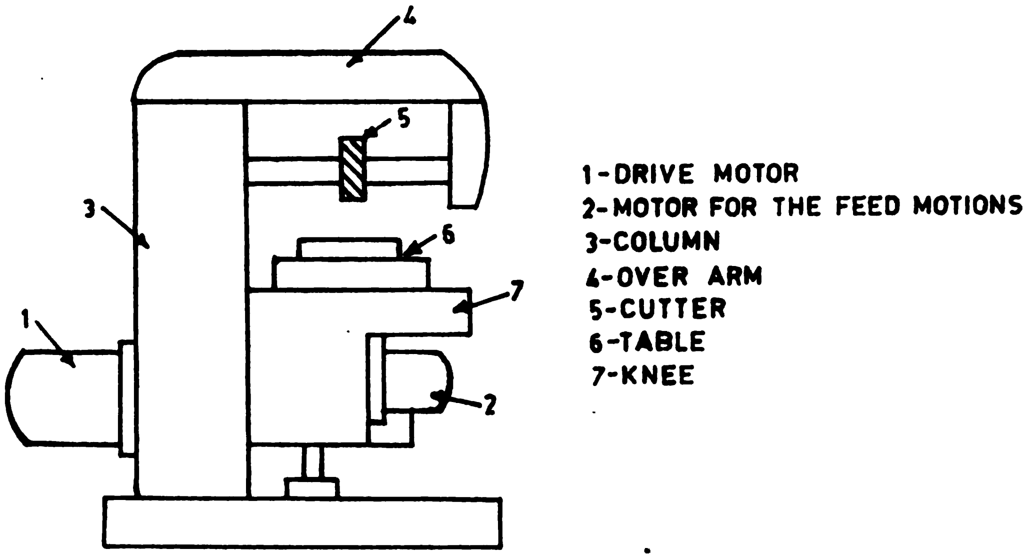


Fig. 8.23 Horizontal milling machine.

The types of jobs performed in a milling machine are:

1. Cutting slots for inserting teeth in milling cutters.
2. Spiral flutes of twist drills.
3. Different types of slots and keyways.
4. Cutting gears.
5. Cutting teeth in racks.
6. Cutting flat surfaces at any angle.

8.7.4 Shaping Machine

A shaping machine (Fig. 8.24) has a reciprocating motion of the ram, which carries the tool. The job is clamped on the table, which is fed horizontally crosswise. The ram has a quick return motion, which is achieved by a crank and stolled arm mechanism. The power is transmitted from a motor by a belt drive through a changed speed gearbox to a bull gear, which carries the crankpin engaging into the stol of a rocker arm. The feed motion is taken from the main drive and the table is cross-fed by an adjustable crankrocker mechanism. The purpose of the shaping machine is to produce flat surfaces. In the shaper, the work is stationary and the tool moves over it. The drive motor is either a three-phase induction motor or a pole-changing induction motor. The speed reduction takes place through belt and gears.

8.7.5 Planer

In a planer (Fig. 8.25), the table mounting the work-piece has a long reciprocating movement. The table is driven by a three-phase induction motor/variable speed dc motor/hydraulic motor though

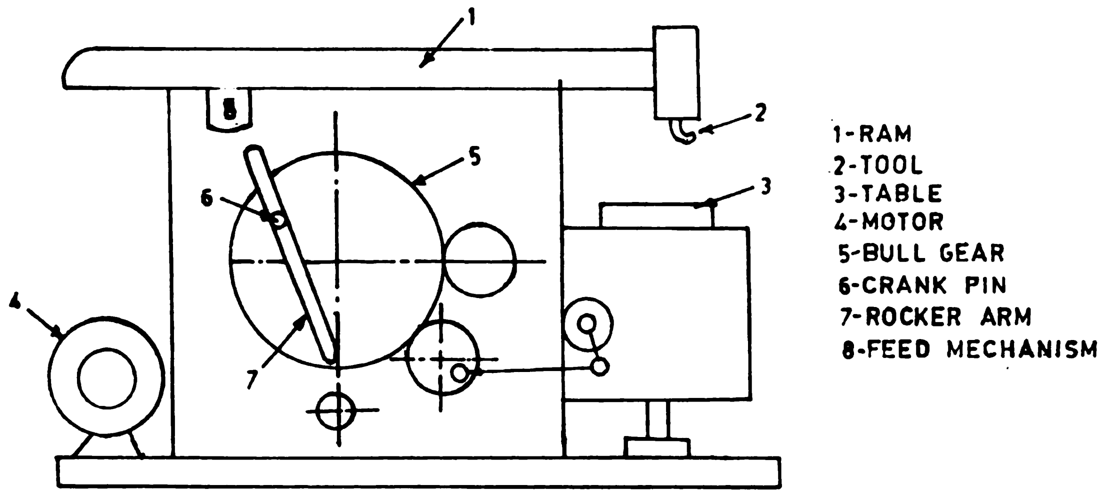


Fig. 8.24 Shaping machine.

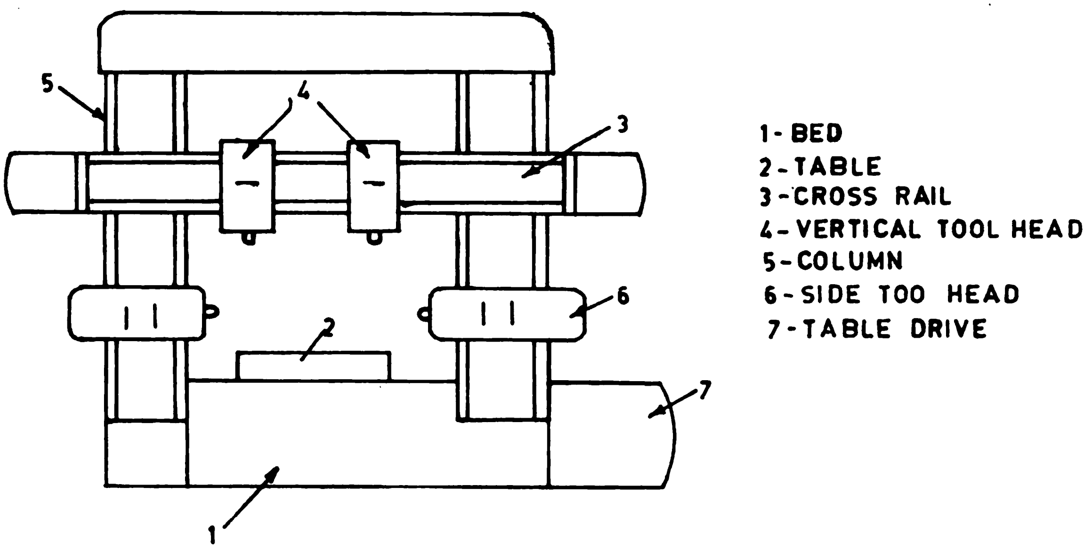


Fig. 8.25 Planer machine.

a gearbox. The reciprocating movement of the feed table may be obtained by a straight and cross-belt pulley. The machine may have one or two vertical tool heads placed on the cross rail. They are driven by the main drive motor for traverse in the horizontal direction. The tools can also be moved vertically on sides of the tool head.

The cross rail can also be lowered or elevated by a motor. There are also side tool heads, which can travel vertically along the column guideways, while their sides carrying the tool can travel horizontally. Several numbers of motors are used as needed for independent operations of tool heads. Alternatively, a single motor can be used coordinating the movement of the tool heads and the to and fro motion of the table.

8.7.6 Grinding Machine

In a grinding machine, the cutter is a grinding wheel for machining the work-pieces with a fine surface finish that is achieved by stepless speed control. Three-phase induction motors are used in association with reduction gears and belt pulleys. The principal movement is the high rotational speed of the grinding wheel. The hydraulic and electromagnetic controls may be used for smooth speed control. The PWM inverter-fed induction motors may also be used for this purpose. In a cylindrical grinder, there is a three-phase SCIM motor for feed motions, which are slow rotation of the work-piece and longitudinal movement of the work table. The wheel head and the work table can be set at an angle for grinding a taper.

8.7.7 Hydraulic Drives

Hydraulic drives are now more widely used as principal and feed movement drives in modern machine tools. The advantages of hydraulic systems are less floor space, capability of infinitely variable speed over a wide range, smooth reversal, automatic overload protection, etc. The shortcomings are leakage of hydraulic fluid, which have to be minimized by very careful manufacture of the components. Also, the properties of hydraulic fluids are affected by variations in temperature. There are different types of pumps, and pressure and flow control devices for use in hydraulic drives.

8.8 TEXTILE MILL

In a composite textile mill, the production of cloth takes place in two stages. The first stage involves a spinning mill for transformation of raw cotton to finished yarn. In the second stage, weaving, dyeing and printing of cloth are done.

8.8.1 Spinning Mill

The following operations are undertaken in a spinning mill:

1. **Picking.** Raw cotton is processed in a blow room to form laps, which resemble rolls of absorbent cotton. In the initial process, the raw cotton is dumped on a conveyor belt feeding the blender, which breaks the raw cotton and separates out the heavier impurities such as seeds and dirt. The cotton is then further beaten and combed to pick out the dirt and put the fibres uniformly along the length of the laps, which are formed by rolling at the finishing end of the blow room. A number of cage induction motors are used for sequential operations of the machines in the blow room. Individual drives are preferred to a group drive. The required speeds of the different parts of the machine are obtained through gears/belts. The ratings of the cage induction motors used range from 0.37 to 7.5 kW for a moderate blow room.

2. **Carding.** From the blow room, the lap is fed to the carding machine in which the cotton fibres are formed by combing them over a revolving cylinder provided with teeth. In this machine the fibres are cleaned as well. All short lengths are removed. The lap is transformed into a wide and thin web of cotton. The web is then gathered together to form a rope, which is known as sliver and

gathered in cylindrical cones. The carding machine may be group driven or individual motor driven. The commonly used carding motors are three-phase cage induction motors of ratings between 1.1 to 3.7 kW. High starting torque is required to overcome the inertia torque of the cylinder. Belts/gears are used to change speed.

3. Drawing. From the carding machines, the slivers pass through the drawing operation, which involves two stages, i.e. preliminary and finisher. Sixteen slivers are combined to make two thinner slivers in the preliminary stage. In the finisher stage, 16 slivers of the preliminary stage are again combined to make two finishers or pucca slivers. In both stages, dust collection is done and fibres remain parallel. The number of stages depends upon the fineness of the yarn required. Three-phase cage induction motors of ratings ranging between 3.7 to 5.6 kW are used for drawing operation. Another 1.5 to 2.2 kW induction motor is used for dust collection. The speeds required for various parts of the machine are obtained through gears/belts.

4. Roving. Twisting process takes place in roving operation. The pucca slivers move through slubbing frames and are wound on a slubbing bobbin giving a few twists. The twist per inch given on roving depends on mill practice. During the roving operation, the tension is adjusted. Six-pole, 7.5 kW, three-phase cage induction motors are commonly used. The motor is provided with starting, inching and stopping controls. A roving machine is also called the *simplex machine*.

5. Spinning. The slubbing bobbins from the roving frame (simplex machine) are fitted to the spindles of the spinning frame (ring frame). The spinning frame continues the process of drawing out and twisting until the finished yarn is produced, which is then wound on smaller bobbins. The spinning frame has a large number of spindles (e.g. 420). A single speed four-pole or six-pole cage induction motor rated for 7.5 to 11.3 kW is normally used. The starting torque should be moderate so that smooth acceleration takes place without tearing the thread. Tension may be adjusted.

8.8.2 Weaving

The weaving process consists of interlacing two sets of yarn at right angles. The lengthwise yarn or threads, which are called warp, interlace with crosswise threads, which are called weft or filling. Individually, the threads are called picks. A loom is a machine for weaving cloth by interlacing warp and filling threads according to a prearranged plan and design. It may be hand or power driven. There are six chief motions in a loom, of which the principal motions are shedding, picking and beating-up and the auxiliary motions are taken-up, let-off and filling pattern.

Principal motions

1. Shedding motion. Warp threads are put in two layers—upper and lower, creating a space between them. This open space is called the shed, through which a shuttle passes back and forth carrying filling yarn. This shed is produced with each turn of the crankshaft of the loom. There is a programming device, which controls the colours of various filling threads.

2. Picking motion. This is the motion of the shuttle through the shed of the loom. The shuttle passes below the raised ends and above the lowered ends.

3. Beating-up. The loose picks are beaten into shed to produce a compact cloth. It happens with turning of the crankshaft.

Auxiliary motions

4. Take-up motion. This is related to the movement of the sand roller with respect to the take-up wheel. The cloth roller is controlled by the take-up motion.

5. Let-off motion. The unwinding of the warp threads from the warp cylinder is known as let-off.

6. Filling pattern motion. This motion is required, where more than one colour is used.

8.8.3 Loom Motors

In power looms, six- or eight-pole, three-phase cage induction motors are used. These motors are normally dripproof, totally enclosed, fan-cooled type. These are subjected to frequent starts and stops. Torque and current pulsations occur due to reciprocating motion of the shuttle. As a result, the temperature rise of the motor is high. High starting torque is required. A flywheel is used to obtain quick pick-up and reduce power consumption. The ratings of the loom motors used for light fabrics like cotton, silk, etc. range between 0.37 to 1.5 kW, while for heavy fabrics the ratings range from 2.2 to 3.7 kW.

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ANSWERS TO PROBLEMS

2.1 (a) 1.006 Ω, (b) 1.424 Ω, (c) 1.98 Ω

2.2 (a) 54 Ω, (b) 50 Ω

2.3 1249 rpm

2.4 (a) 235.75 A, (b) 75%

2.5 *Natural speed characteristic*

I (A)	8.0	12.0	16.0	20.0	24.0	28.0	32.0	36.0	40.0
N (rpm)	2000	1300	1090	1000	900	820	780	720	680

External resistances
2.023 Ω, 4.046 Ω

Rheostatic characteristics

I (A)	8	12	16	20	24	28	32	36	40
N_{nat} (rpm)	2000	1300	1090	1000	900	820	780	720	680
N_{rh1} (rpm)	1859	1157	923	800	673	566	489	400	324
N_{rh2} (rpm)	1719	1015	756	600	446	312	198	81	−32

2.6 6.6 Ω

2.7 3.612 W, 797 N · m, 606.7 N · m

2.8 25.22%

2.9 57.7%, 0.2025

2.10 45.6 N · m, 341.0 N · m

2.11 166.68 N · m

3.1 959 rpm, 102.56 N · m, 93.83 N · m

3.2 (a) 4.15 Ω, (b) 1.383 Ω

3.3 (a) 3.92 Ω, (b) 0.62 Ω

3.4 (a) 245 to 1462 rpm, (b) 175 to 1392 rpm

3.5 (a) 1095 rpm, (b) 177.5 V and 845 rpm, 125 V and 595 rpm, 72.5 V and 345 rpm, (c) Speed range ratio = 1000/190.5 = 5.25

3.6 (a) 617 rpm, 11.3 kW, 81.9%, (b) 3.68 Ω, 11.3 kW, 49.1%

3.7 (a) 6.1 hp, 73.2 N · m, (b) 10.1 Ω, 13.33 A, 1.827 hp

3.8 (a) The circuit diagrams and waveforms are shown in Fig. S3.8.1(a) and (b) and Fig. S3.8.2 respectively. The current (Fig. S3.8.2) in the diode D_1 (Fig. S3.8.1b) also includes the current in the freewheeling path, or the freewheeling diode D_F (Fig. S3.8.1a)

(b) 1.383 kW, (c) 1.8 kVA, 0.768 (lag), (d) 9.186 A, 9.186 A

3.9 (a) 22.9°, (b) 16.95 A, 12 A, (c) 0.83 (lagging)

The circuit and waveforms are shown in Figs. S3.9.1 and S3.9.2 respectively.

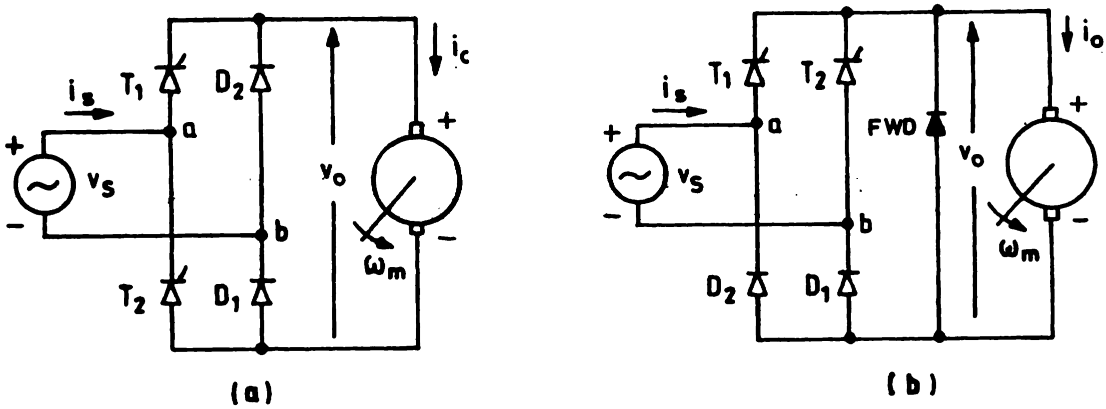

Fig. S3.8.1 Circuit diagrams.

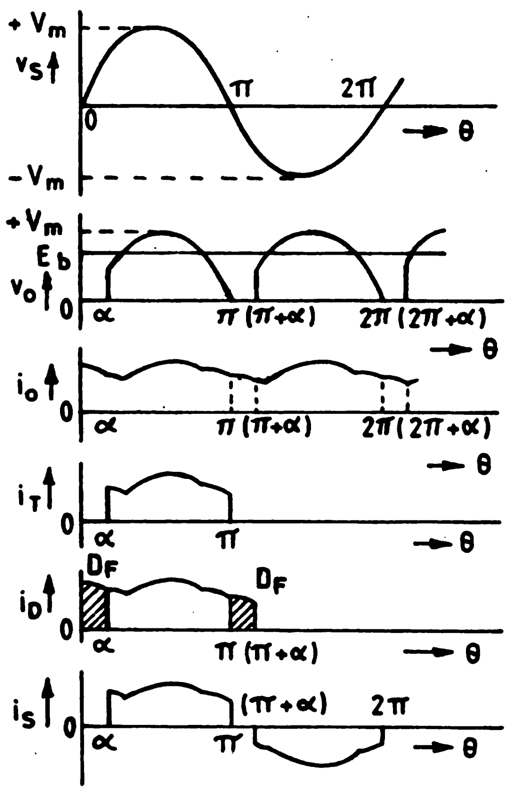

Fig. S3.8.2 Voltage and current waveforms.

3.10 (a) 13 A, (b) 7.54 A, 10.67 A, (c) 0.475 (lagging)

3.11 (a) 0.6, (b) 44 kW, (c) The circuit and waveforms are shown in Figs. S3.11.1 and S3.11.2 respectively.

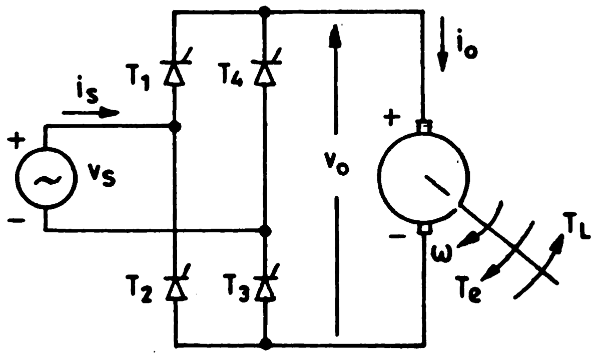

Fig. S3.9.1 Circuit diagram.

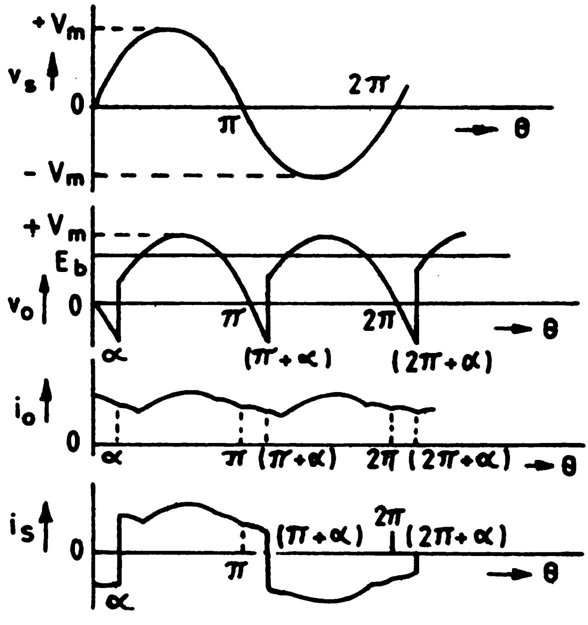

Fig. S3.9.2 Voltage and current waveforms.

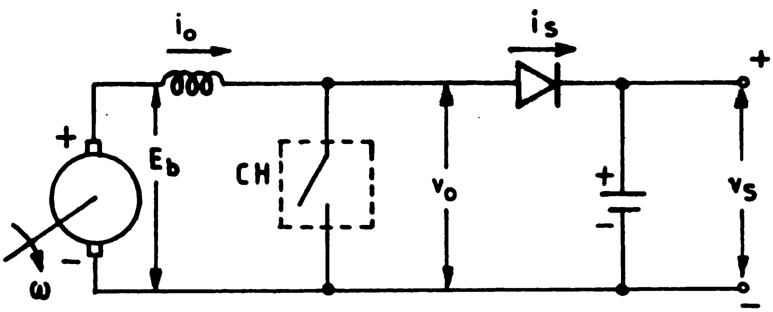

Fig. S3.11.1 Circuit diagram.

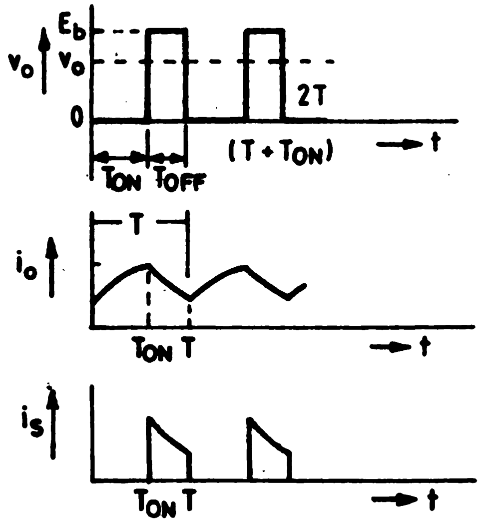

Fig. S3.11.2 Voltage and current waveforms.

4.1 1.7 Ω

4.2 6.77 Ω

4.3 (a) 525 or 975 rpm (for f = 15 Hz), – 750 or 2250 rpm (for f = 100 Hz)
(b) 51 V, 340 V, (c) For 15 Hz, $P_{ac} = 3.333\ P_2$, $P_{dc} = -2.333\ P_2$, For 100 Hz, $P_{ac} = -0.5\ P_2$, $P_{dc} = 1.5\ P_2$, where P_2 = rotor power output.

4.4 214 rpm

4.5 429, 3000, 750, 1000 rpm

4.6 0.63 Ω

4.7 – 38.6 V ∠5.7°

4.8 (a) 0.608, 0.033, (b) 588 rpm

4.9 (a) –108.77 N · m, 1500 ≤ N ≤ 1985 (rpm), (b) 1639 rpm, (c) 1.69

5.1 25.82 kW

5.2 17 kW

5.3 405.8 kW

5.4 $x = \sqrt{\dfrac{(1+\alpha)\{1-e^{-t_1/\epsilon T}\}}{1-e^{-t_1/T}} - \alpha}$

5.5 Half-hour rating is 2.243 times the continuous rating

5.6 (a) 92.5°C, (b) 58.73 min, (c) 82.22 min

5.7 (a) 74 kW, (b) 60.6 kW

5.8 290.8 kW

5.9 (a) 54.8 kW, (b) 28.3 kW

5.10 646 kg · m²

6.1 105.37 kg · m², 1.58 kW

6.2 (a) $\omega_m(t) = 111.86\,(1 - e^{-t/0.9887})$ rad/s, (b) 1068 rpm, (c) 1.37 sec

6.3 Natural frequency ω_n = 32.635 rad/s
Damping ratio ζ = 1.9407

$$\Omega(s) = \frac{11.82}{s} - \frac{12.806}{s + 9.0557} + \frac{0.986}{s + 117.61}$$

$\omega(t)$ (exact) = $11.82 - 12.806\, e^{-9.0557t} + 0.986\, e^{-117.61t}$ rad/s
$\omega(t)$ (approximate) = $11.82(1 - e^{-8.408t})$ rad/s

6.4 $T_e = T_{\text{nom}} + J\dfrac{d\omega_m}{dt}$

$$3.42\frac{d\omega_m}{dt} + \omega_m = 38.23$$

$\tau_{em} = 3.42$ sec
$N_m = 365$ rpm

6.5 22.44 $\text{kg} \cdot \text{m}^2$

6.6 0.445 sec

6.7 (a) 8.57 Wh (one step), 4.28 Wh (two steps),
(b) 0.0258 kWh,
(c) 0.00214 kWh (regenerative braking), 0.00643 kWh (counter current braking)

6.8 4.33 sec, 1.27 Ω

6.9 1500 rpm

6.10 0.312 sec

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ELECTRIC DRIVES

N.K. DE · P.K. SEN

This book provides a comprehensive introduction to the fundamental concepts of electric drives and is eminently suited as a textbook for B.E./B.Tech., AMIE and diploma courses in electrical engineering. It can also be used most effectively by all those preparing for GATE and UPSC competitive examinations, as well as by practising engineers. The topics, which range from principles and techniques to industrial applications, include characteristic features of drives, methods of braking and speed control, electromagnetic and solid state control of motors, motor ratings, transients in drive systems, and operation of stepper motors.

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