Chapter 3

Speed control of an Induction motor

3.1 INTRODUCTION

Apart from the nonlinear characteristics of the induction motor, there are various issues attached to the driving of the motor. Let us look at them one by one. Earlier motors tended to be over designed to drive a specific load over its entire range. This resulted in a highly inefficient driving system, as a significant part of the input power was not doing any useful work. Most of the time, the generated motor torque was more than the required load torque.

For the induction motor, the steady state motoring region is restricted from 80% of the rated speed to 100% of the rated speed due to the fixed supply frequency and the number of poles. When an induction motor starts, it will draw very high inrush current due to the absence of the back EMF at start. This results in higher power loss in the transmission line and also in the rotor, which will eventually heat up and may fail due to insulation failure. The high inrush current may cause the voltage to dip in the supply line, which may affect the performance of other utility equipment connected on the same supply line. When the motor is operated at a minimum load (i.e., open shaft), the current drawn by the motor is primarily the magnetizing current and is almost purely inductive. As a result, the PF is very low, typically as low as 0.1. When the load is increased, the working current begins to rise. The magnetizing current remains almost constant over the entire operating range, from no load to full load. Hence, with the increase in the load, the PF will improve. When the motor operates at a PF less than unity, the current drawn by the motor is not sinusoidal in nature. This condition degrades the power quality of the supply line and may affect performances of other utility equipment connected on the same line. The PF is very important as many distribution companies have started imposing penalties on the customer drawing power at a value less than the set limit of the PF. This means the customer is forced to maintain the full-load condition for the entire operating time or else pay penalties for the light load condition. While operating, it is often necessary to stop the motor quickly and also reverse it. In applications like cranes or hoists, the torque of the drive motor may have to be controlled so that the load does not have any undesirable acceleration (e.g., in the case of lowering of loads under the influence of gravity). The speed and accuracy of stopping or reversing operations improve the productivity of the system and the quality of the product. For the previously mentioned applications, braking is required. Earlier, mechanical brakes were in use. The frictional force between the rotating parts and the brake drums provided the required braking. However, this type of braking is highly inefficient. The heat generated while braking represents loss of energy. Also, mechanical brakes require regular maintenance.

In many applications, the input power is a function of the speed like fan, blower, pump and so on. In these types of loads, the torque is proportional to the square of the speed and the power is proportional to the cube of speed. Variable speed, depending upon the load requirement, provides significant energy saving. A reduction of 20% in the operating speed of the motor from its rated speed will result in an almost 50% reduction in the input power to the motor. This is not possible in a system where the motor is directly connected to the supply line. In many flow control applications, a mechanical throttling device is used to limit the flow. Although this is an effective means of control, it wastes energy because of the high losses and reduces the life of the motor valve due to generated heat.

When the supply line is delivering the power at a PF less than unity, the motor draws current rich in harmonics. This results in higher rotor loss affecting the motor life. The torque generated by the motor will be pulsating in nature due to harmonics. At high, speed, the pulsating torque frequency is large enough to be filtered out by motor impedance. But at low speed, the pulsating torque results in the motor speed pulsation. This results in jerky motion and affects the bearings life. The supply line may experience a surge or sag due to the operation of other equipment on the same line. If the motor is not protected from such conditions, it will be subjected to higher stress than designed for, which ultimately may lead to its premature failure.

All of the previously mentioned problems, faced by both consumers and the industry, strongly advocated the need for an intelligent motor control. With the advancement of solid state device technology (BJT, MOSFET, IGBT, SCR, etc.) and IC fabrication technology, which gave rise to high-speed microcontrollers capable of executing real-time complex algorithm to give excellent dynamic performance of the AC induction motor, the electrical Variable Frequency Drive became popular.

The speed of an induction motor is given by

$$N = \frac{120 f}{P}(1 - s)\_\_\_\_\_\_(3.1)$$

From above equation, it is clear that we can change the speed of the motor by either changing frequency of the supply voltage, number of poles of the winding, or slip of the motor.

The methods of changing the speed of an induction motor can be divided in to two parts. (1) Control from stator side (1-a) Changing supply voltage, (1-b) changing frequency of supply voltage (1-c) changing number of poles of the stator winding (2) control from rotor sides (2-a) changing resistance in rotor winding, (2-b) injecting emf in rotor circuit.

The methods of controlling speed from stator sides are applicable to both the motors viz. slip ring motor and squirrel cage motor, where as methods of controlling speed from rotor side are applicable to only slip ring motor.

Stator Voltage Control of an Induction Motor is used generally for three purposes (a) to control the speed of the motor (b) to control the starting and braking behaviour of the motor (c) to maintain optimum efficiency in the motor when the motor load varies over a large range. It is simple in hardware and reliable compared to the more complex Variable Frequency Drives as far as speed control application is concerned. However, it turns out to be a somewhat dissipative method of speed control and results in lowered efficiency and rotor overheating. Fundamental aspects of Stator Voltage Control aimed at the above three objectives is covered in this lecture.

3.2 STATOR VOLTAGE CONTROL FOR SPEED CONTROL OF INDUCTION MOTOR:

With fixed frequency and variable magnitude pure sine wave source the torque at any particular slip is proportional to square of Voltage in an Induction Motor. Fig. 3.1 below shows the Torque-Speed curves of 3-phase, 5 HP, 4-pole, 415 V, 50 Hz Induction Motor at various sinusoidal voltages. Also included is the torque-speed curve of a typical fan load. Note that the torque-speed curves of the Induction Motor clearly indicate that it is a motor designed for a large running slip. Otherwise the curves would have been steeper than this around full load rated speed. It can be clearly seen that the speed of the fan can be varied more or less uniformly in the range of 90% to 40% of synchronous speed of the motor by varying the voltage between 100% and 40%.

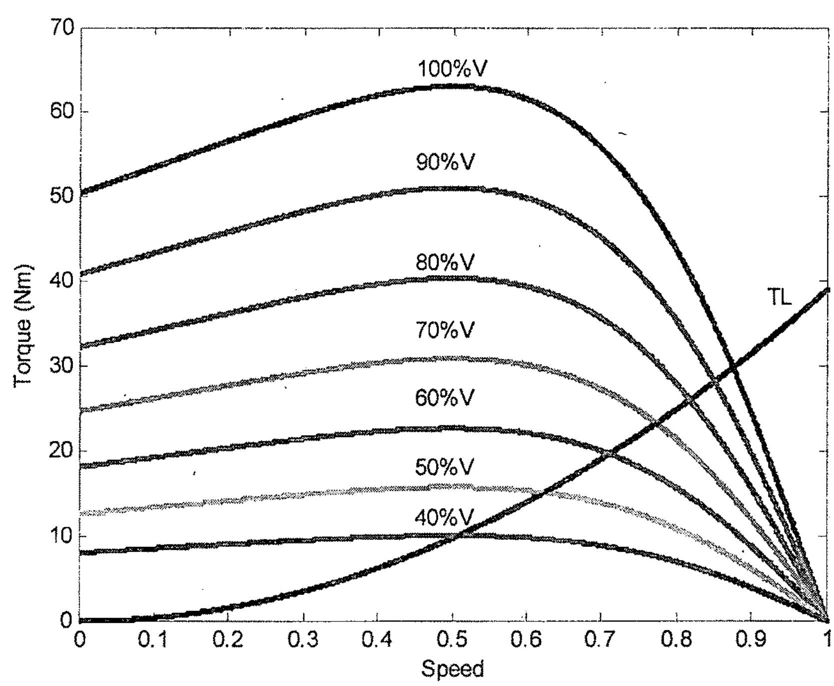


Fig. 3.1 Torque speed curve of fan motor for various voltages.

Had the torque-speed curves of the Induction Motor been steep around synchronous speed (as it is in the case of a well designed Squirrel cage Motor with a running slip in the range of 2% to 5%) the possible speed range in the case of a fan load with voltage control would have been lower than this. The motor will pull out at around 55% voltage or so. Moreover, even the available speed variation will be highly non-uniform with the voltage variation. Hence it is necessary that an Induction Motor intended for speed control applications using stator voltage control has to be designed with a higher running slip and hence lower full load efficiency. (Because higher full-load slip implies higher rotor resistance). Such motors are usually designed with a full load slip value of 12%. Obviously their rotor must be of special design in order to withstand the higher rotor losses developed in the rotor.

The range of speed control available by voltage control is a strong function of nature of load torque variation with load. With a constant torque load, the available speed range is more limited than in the case of a fan load and the motor pulls out at voltage levels closer to 100%. Thus fan loads which have low starting torque demand, pump loads with little or no static head component in the system curve, blower loads with small starting torque demand etc. are the loads suitable for speed control by voltage variation.

3.2.1 VARIATION OF STATOR CURRENT AND EFFICIENCY IN STATOR VOLTAGE CONTROL

For simplicity the analysis to follow will neglect the magnetising current of the machine. Though the magnetising current can be as much as 50% of full load current at rated voltage, it comes down rapidly with voltage and hence the above assumption is reasonable.

Assuming sinusoidal quantities, the average torque produced by the stator field reacting with rotor current is given by the following proportionality:

$$T_m \propto \frac{I_2^2 * R_2}{s} \qquad (3.2)$$

where T_m is the motor torque, R_2 the rotor resistance, I_2 the rotor current and s is the slip.

Neglecting the magnetising current and core loss current the stator current is proportional to rotor current. If the load torque is related to the square of motor speed as is approximately true for a fan load, then

$$T_L \propto (1-s)^2 \qquad (3.3)$$

where T_L is the load torque.

At steady state the motor torque and load torque will be equal. This results in the following proportionality for stator current.

$$I_1 \propto \frac{(1-s)\sqrt{s}}{\sqrt{R_2}} \qquad (3.4)$$

This function has a maximum at $s = 1/3$ (for fix value of rotor resistance differentiate numerator and equate to zero).

Thus *for a true fan type load stator voltage control will result in a maximum current at 66.7% of synchronous speed.*

The ratio of this maximum current at s = 0.33 to the rated full load current of the motor will vary sharply with the rated full load slip. The ratio is 1.75 if full load slip is 0.05 and it is 1.25 if the full load slip is 0.12.This implies that the maximum rotor copper loss (and stator Copper loss) in the machine will take place when the voltage applied is such that the fan load runs with a slip of 0.33 and that this maximum loss will be closer (but higher) to the rated full load copper loss if the rated slip of the machine is higher than normal (i.e. around 5%). Usually the motors for this service are designed with a full load slip of 0.12 and hence their current can go up to 25% higher than rated value as the speed of a fan load is varied by varying voltage.

Similarly their copper losses can go to 50% more than the full load copper loss under a variable voltage-fan load context. Thus we need an inherently inefficient machine to start with and the machine operation gets more and more inefficient at lower speeds. Hence this kind of speed control is used only on fan type loads and when only about 60% to 100% speed range is needed. Even then a motor with high rotor resistance (i.e. with a running slip of about 12%) should be used. Ordinary Squirrel Cage motors will suffer from rotor over heating on stator voltage control and should not be used for such service. Substituting s = 1/3 in equation (3.4) we have maximum current equation as

$$I_m \propto \frac{(1-1/3)\sqrt{1/3}}{\sqrt{R_2}} \text{\_\_\_\_\_\_\_\_}(3.4)$$

And rated current as

$$I_R \propto \frac{(1-s_R)\sqrt{s_R}}{\sqrt{R_2}} \text{\_\_\_\_\_\_\_\_}(3.5)$$

Taking ration of maximum rotor copper loss to rated copper loss, we have

$$\frac{Maximum\ rotor\ copper\ loss}{Rated\ copper\ loss} = \frac{\frac{4}{27}}{s_R(1-s_R)^2} \text{\_\_\_\_\_\_\_\_}(3.6)$$

Variation of maximum rotor copper loss with rated slip is shown in fig. 3.2

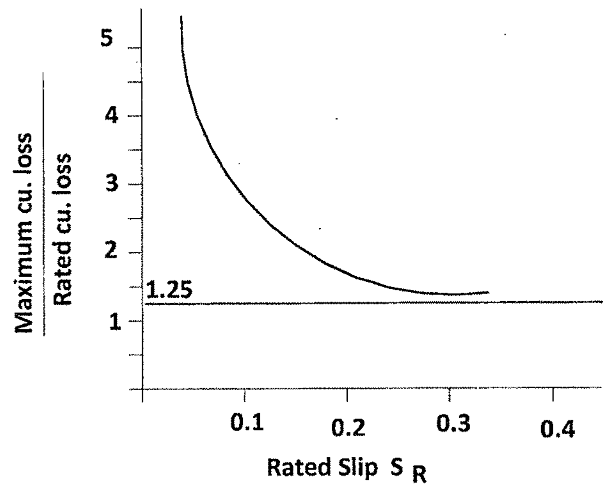


Fig. 3.2 Variation of ratio of maximum cu losses to rated cu loss for Induction motor fan type load

3.2.2 THYRISTOR BASED STATOR VOLTAGE CONTROLLER

The stator voltage is controlled in these speed control systems by means of a power electronic controller. Normally thyristors in phase control mode are used. Various connection schemes exist. However detailed investigations into various connections had established in early eighties that the six thyristor-unconnected neutral scheme is the best in terms of minimum r.m.s current requirement and harmonic injection. Here two thyristors in anti parallel are connected between the line and motor in a phase as shown in fig. 3.3. Typical wave form of voltage and current for this method is shown in fig. 3.4. If the motor is star connected the neutral is left unconnected. This scheme was proved to take only 8% more r.m.s current (due to harmonics and converter induced reactive power requirement) than a pure sine wave source at the maximum current slip value of s = 0.33. All other possible thyristor connections take more than this. Hence this 6-thyristor scheme is almost invariably used to control the applied voltage to Induction Motors in speed control schemes. The control is exercised by changing the firing angle alpha of thyristors.

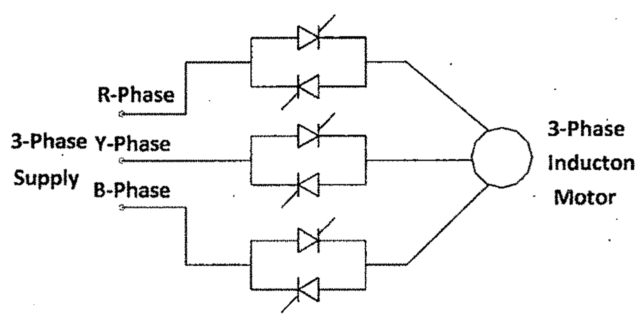


Fig. 3.3 Thyristor Voltage controller for Induction Motor

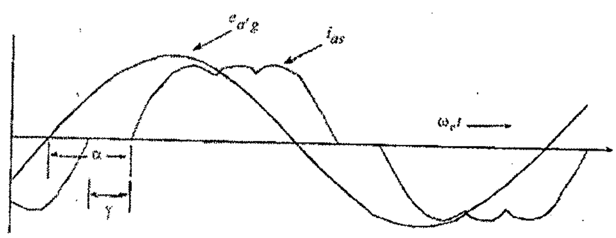


Fig.3.4 Wave form of voltage and current with thyristor voltage control.

The thyristor controller brings in two more sources of power loss. Power loss takes place in the power devices in the controller. In addition, harmonic losses take place in the motor due to harmonic currents flowing in the winding due to phase control. These two additional loss components will make this speed controller further inefficient. Also over heating of the motor on harmonic losses is another possibility. Harmonic currents can result in cogging/crawling etc. especially when attempts are made to run the motor at very low speeds. In spite of all these problems these speed controllers are popular; especially for fan loads with limited speed variation, due to their simplicity and reliability.

3.2.3 STARTING/STOPPING CONTROL BY STATOR VOLTAGE CONTROL

The commercially available Starting Torque Controllers (STC), Soft Starters etc. make use of stator voltage control using the 6-thyristor scheme to control the starting/stopping behaviour of the motors. They behave like a continuously variable autotransformer during starting. After starting, the thyristors are shorted by contactors to avoid device losses and full voltage is directly applied to the motor.

3.2.4 STATOR VOLTAGE CONTROL FOR OPTIMUM EFFICIENCY OPERATION OF MOTOR

Induction Motors are highly efficient at rated load and have efficiencies in the range 85%-95%. Motor losses consist of three main components: (1) Friction and Windage Losses (2) Iron Losses (3) Copper Losses. Friction and Windage Losses are insensitive to load changes, as speed is essentially constant.

Iron losses consist of hysteresis losses and eddy current losses. At constant frequency, hysteresis loss is proportional to $\boldsymbol{B}^{1.6}$ and the eddy current loss is proportional to $\boldsymbol{B}^2$, where $\boldsymbol{B}$ is the maximum flux density in the air gap. The maximum flux density remains constant if the applied voltage is kept constant. Thus, as load is decreased, voltage remaining constant, the iron loss constitutes a greater percentage of the output. This results in poor efficiency at part loads.

Part load efficiency can be improved by reducing the applied voltage to the motor. The motor has to be a standard squirrel cage motor optimised for full load running. In the case of such a motor the running slip will be around 0.04 and hence its torque-slip curve will be steep around zero slip. When the applied voltage is reduced, the load torque intersects the motor curve at a new point on the new torque-slip curve. However due to the steepness of T-s curves, the speed of machine will not vary much though it will decrease a little. Hence as a first approximation it may be assumed that the motor speed does not change when voltage across an under loaded motor is varied. If the speed does not change the load torque and mechanical power output will not change. And since voltage has come down the motor will draw an increased active current component to supply the same output. The reactive current component is predominantly magnetising in nature and it will come down since applied voltage has come down. The total stator current which is constituted by active and reactive components can increase or decrease depending on the amount of voltage reduction. Thus when the voltage across an under loaded motor is gradually reduced its stator current decreases first, reaches a minimum at a particular voltage and increases with further reduction in voltage. The value of minimum current will depend on the exact load on the motor.

Coming to the loss variations, with reduction in voltage the iron loss comes down. And initially the currents and hence copper losses also come down. When the voltage is reduced to sufficiently low level, the consequent increase in copper losses will at some point turn the total losses away from its decreasing trend i.e. there will be one particular voltage at which the total losses in the motor will be a minimum. This voltage value will not coincide with the voltage value at which the current is a minimum at the same loading level; but they will be close.

With the assumption that the speed of the motor does not vary with reduction in voltage the minimum current point will coincide with maximum power factor (or minimum phase angle) condition. Similarly the minimum loss point (i.e. maximum efficiency point) will coincide with minimum power input point. Minimum current point does not correspond to maximum efficiency point as already mentioned; but they are close. But if the small variation in motor speed and consequent changes in output power are also considered, the optimum voltage point for a particular load condition in the four cases i.e. the minimum current point, the minimum power factor angle point, the minimum power input point and the minimum loss (maximum efficiency) point, will be different. The minimum loss point is difficult to monitor electronically; though that is what we want to do. However, the other three conditions can be monitored electronically by sensing motor voltage and current and using some form of a minimum search algorithm implemented either digitally or in analog circuits. Of course, the loss reduction achieved will be less than optimal. Minimum power condition is the closest to maximum efficiency condition followed closely by current minimum condition. It is easier to process the current minimum search and hence it is current minimum search that is employed in most of the Smart Motor Controllers available in the market.

The six-thyristor scheme is used in all these SMCs. The SMCs also take care of the control of starting and stopping of the motor also. Essentially, they start up the motor and apply full voltage first. Then, the current is sampled. A search routine is initiated. The voltage is decreased slightly and the change in current is noted. If the current decreases the voltage is further reduced in steps till the current shows a tendency to turn back i.e. to increase. If, in the first voltage reduction step the current increased, then the voltage is taken up in steps till the current reaches a minimum. This procedure is repeated in a periodic manner to fine-tune the applied voltage against load variations.

3.2.5 BASIC PRINCIPLES OF VOLTAGE CONTROL

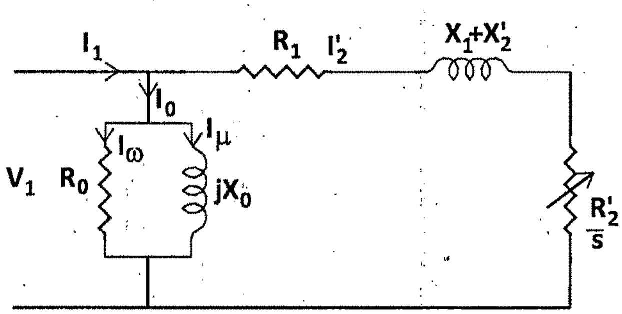


Fig. 3.5 Simplified Approximate equivalent circuit of an Induction Motor referred to stator

The basic principles of voltage control can be obtained readily from the conventional induction motor equivalent circuit shown in Figure 3.5 and the associated constant voltage speed-torque curves illustrated in Figure 3.6. The torque produced by the machine is equal to the power transferred across the air gap divided by synchronous speed,

$$T = \frac{3}{2} * P\left(\frac{I_2^2 * R_2}{s * f}\right) \text{\_\_\_\_\_\_\_\_\_\_\_\_\_\_}(3.7)$$

where P = number of poles, s is the per unit slip, f is line frequency and I_2, R_2 are the rotor rms current and rotor resistance respectively. Approximately this equation can be written as

$$T = \frac{3}{\omega_s * s}\left(\frac{V_1^2 * R_2'}{\left(R_1 + \frac{R_2'}{s}\right)^2 + (X_1 + X_2')^2}\right) \text{\_\_\_\_\_\_\_\_\_\_\_\_\_\_}(3.8)$$

The peak torque points on the curves in Figure 3.7 occur when maximum power is transferred across the air gap and are easily shown to take place at a

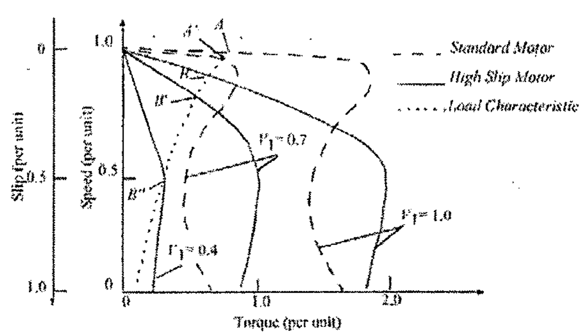


Fig.3.6 Torque slips Curves for standard and high slip Induction Motor.

Slip

$$s_m = \pm \frac{R_2'}{R_1 \pm \sqrt{(R_1^2 + (X_1 + X_2')^2)}} \text{\_\_\_\_\_\_\_\_}(3.9)$$

Substituting equation (3.9) into equation (3.8) gives the equation of maximum torque as

$$T_m = \frac{3}{2 * \omega_s}\left(\frac{V_1^2}{R_1 \pm \sqrt{(R_1^2 + (X_1 + X_2')^2)}}\right) \text{\_\_\_\_\_\_\_\_\_\_\_\_\_\_}(3.10)$$

where X_1 *and* X_2' are the stator and rotor leakage reactance. From these results and the equivalent circuit, the following principles of voltage control are evident.

(1) For any fixed slip or speed, the current varies directly with voltage and the torque and power with voltage squared.

(2) As a result of (1) the torque-speed curve for a reduced voltage maintains its

shape exactly but has reduced torque at all speeds, see Figure 3.6.

(3) For a given load characteristic, a reduction in voltage will produce an increase in slip (from A to A' for the conventional machine in Figure 3.6,

(4) A high-slip machine has relatively higher rotor resistance and results in a larger speed change for a given voltage reduction and load characteristic. (compare A to A' with B to B' in Figure 3.6).

(5) At small values of torque, the slip is small and the major power loss is the core loss inR_0. Reducing the voltage will reduce the core loss at the expense of higher slip and increased rotor and stator loss. Thus there is an optimal slip which maximizes the efficiency and varying the voltage can maintain high efficiency even at low torque loads.

It has been shown that a very accurate fundamental component model for a voltage converter comprised of inverse parallel thyristors (or Triacs) is a series reactance given by [19]:

$$x_{eq} = x_s{}' * f(\gamma) \qquad (3.11)$$

where

$$x_s' = x_1 + \frac{x_0 * x_2{}'}{x_0 + x_2{}'} \; and \; x_1, x_2{}', x_0 \qquad (3.12)$$

are the induction motor stator leakage, rotor leakage and magnetizing reactance respectively and γ is the thyristor *hold-off angle* identified in Figure 3.4 and

$$f(\gamma) = \left(\frac{3}{\pi}\right) \frac{(\gamma + \sin\gamma)}{1 - \frac{3}{\pi}(\gamma + \sin\gamma)} \qquad (3.13)$$

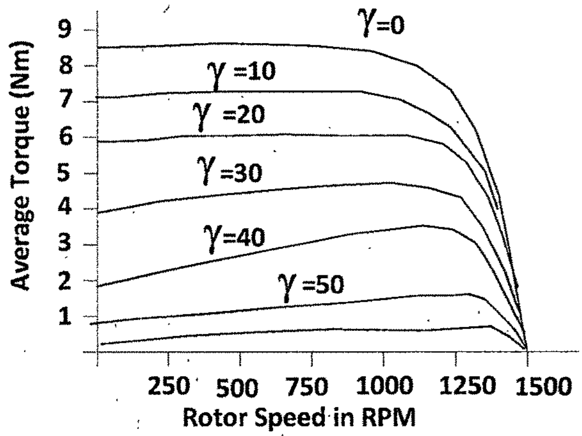


Fig. 3.7 Torque Slip Curves for changing in hold off angle γ

This reactance can be added in series with the motor equivalent circuit to model a voltage-controlled system. For typical machines the accuracy is well within acceptable limits although the approximation is better in larger machines and for smaller values of (γ) . In most cases of interest, the error is quite small. However, the harmonic power losses and torque ripple produced by the current harmonics implied in Figure 3.4 are entirely neglected. A plot of typical torque versus speed characteristics as a function of (γ) is shown in Figure 3.7 for a squirrel cage induction machine [20].

3.3 CHANGING FREQUENCY OF SUPPLY VOLTAGE

Synchronous speed and, therefore, the motor speed can be controlled by varying supply frequency. The supply voltage to motor is sinusoidal then it's RMS value is given by equation

$$V = 4.44\,\varphi f\, T_{ph} \_\_\_\_(3.14)$$

Where

$$\varphi = Flux\ per\ pole\ wb, f = frequency\ of\ supply\ voltage, T_{ph} = No. of\ turns\ phase$$

Hence if magnitude of supply voltage is constant product of flux and frequency is constant. For the purpose of changing speed if we decrease the frequency, without a change in the magnitude of voltage, causes an increase in the air-gap flux. Induction motors are designed to operate at the knee point of the magnetization characteristic to make full use of the magnetic material. Therefore, the increase in flux will saturate the magnetic circuit of the motor. This will increase the magnetizing current, distort the line current and voltage, increase the core loss and the stator copper loss, and produce a undesirable noise. Similarly increase in supply frequency will decrease the magnitude of flux and hence reduces the torque capability of the motor. Therefore, the variable frequency control below the rated frequency is generally carried out at rated air gap flux by varying terminal voltage with frequency so as to maintain (V/f) ratio constant at the rated value.

Now substituting value of stator and rotor leakage reactance in terms of inductances in equation (3.10) we have

or

$$T_m = \left(\frac{K\left(\frac{V}{f}\right)^2}{\frac{R_s}{f} \pm \sqrt{\left(\frac{R_s}{f}\right)^2 + 4\pi^2\,(L_s + L_r')^2}} \right) \_\_(3.15)$$

$$Where\ K\ is\ a\ constant = \frac{3P}{8\pi} \_\_(3.16)$$

and L_s and L_r' are, respectively, the stator and stator referred rotor inductances. Positive sign is for motoring operation where as negative sign is for braking operation.

When frequency is large such that

$$\frac{R_s}{f} \ll 2\pi(L_s + L_r') \_\_(3.17)$$

Then from equation (3.1)

$$T_m = \pm\left(\frac{K\left(\frac{V}{f}\right)^2}{2\pi(L_s + L_r')} \right) \_\_(3.18)$$

Equation (3.4) suggests that with a constant *(V/f)* ratio, motor develops a constant maximum torque, except at low speeds (or frequencies). Motor therefore operates in constant torque mode. According to Equation (3.1), for low frequencies (or low speeds) due to stator resistance voltage drop [i.e. when *(R_s/f)* is not operates in constant torque mode. According to Equation (3.1), for low frequencies (or low speeds) due to stator resistance voltage drop [i.e. when *(R_s/f)* is not negligible compared to $2\pi(L_s + L_r')$] the maximum torque will have lower value in motoring (+ve sign) and large value in braking operation (-ve sign). This behaviour is due to reduction in flux during motoring operation and increase in flux during braking operation. When it is require to maintain the same maximum torque at low speeds during motoring operation it is necessary to increase *(V/f)* at low frequencies. This causes further increase in maximum braking torque and considerable saturation of the machine in braking operation.

When voltage reaches rated value corresponding to base speed, it cannot be increased with frequency. Therefore, above base speed, frequency is changed with voltage maintained constant. According to equation (3.18), with voltage maintained constant, maximum torque decreases with increase in frequency.

The variation of voltage and torque with speed is shown in fig. 3.8 and Torque speed curve for *(V/f)* is shown in fig. 3.9

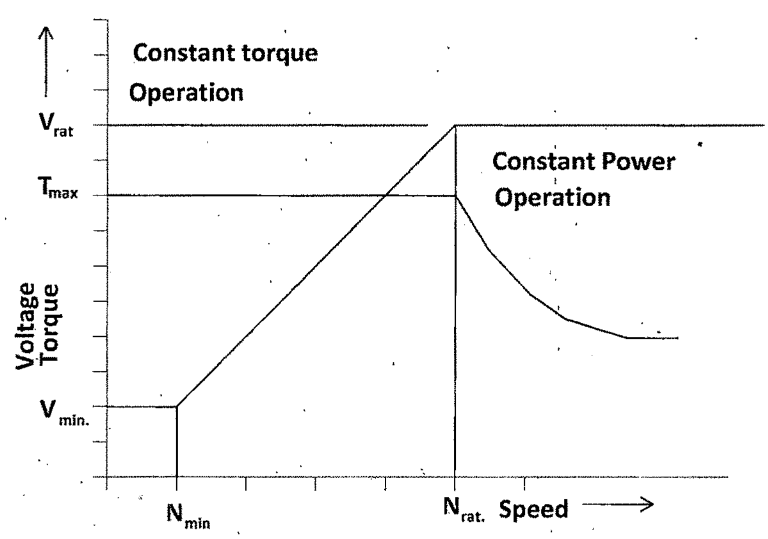


Fig. 3.8 Variation of torque and voltage with speed for V/F control method

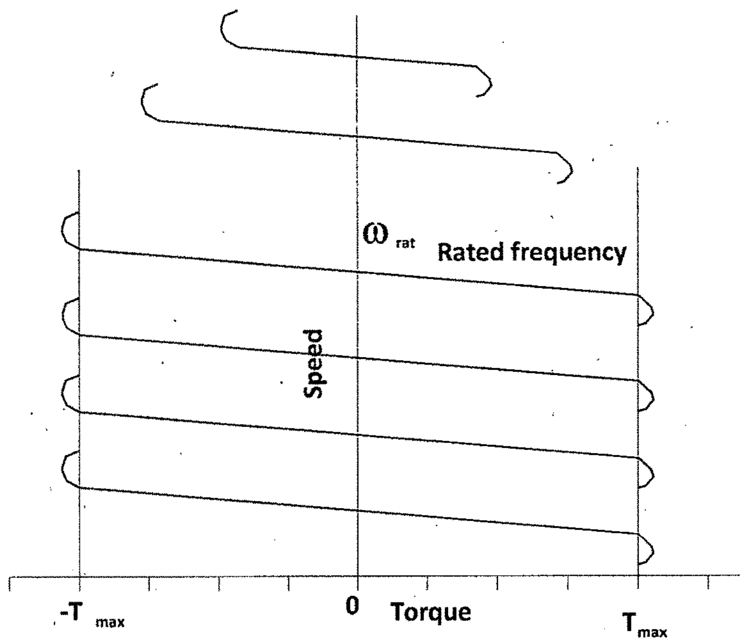


Fig. 3.9 Speed torque curve for V/f control

A given torque is obtained with a lower current when the operation at any frequency is restricted between the synchronous speed and the maximum torque point, both for motoring and braking operations. Therefore, the motor operation for each frequency is restricted between the synchronous speed and the maximum torque point as shown by solid lines in fig. 3.9

The variable frequency control provides good running and transient performance because of the following features:

(i) Speed control and braking operation are available from zero speed to above base speed.

(ii) During transients (starting, braking and speed reversal) the operation can be carried out at the maximum torque with reduced current giving good dynamic response.

(iii) Copper losses are low, and efficiency and power factor are high as the operation is restricted between synchronous speed and maximum torque point at all frequencies.

(iv) Drop in speed from no load to full load is small.

The most important advantage of variable frequency control is this that it allows a variable speed drive with above mentioned good running and transient performance to be obtained from a squirrel cage induction motor. The squirrel cage reliable and long listing. Because of the absence of commutator and brushes, it requires practically no maintenance, it can be operated in an explosive and contaminated environment, and it can be designed for higher speeds, voltage and power ratings. It also has lower inertia, volume and weight. Though the cost of a squirrel cage motor is much lower compared to that of a dc motor of the same rating, the overall cost of variable frequency induction motor drive, in general are higher. But because of the advantages listed above, variable frequency induction motor drives are preferred over dc motor drives for most applications.

In special applications requiring maintenance free operation, such as underground or underwater installation, and also in applications involving explosive and contaminated environments, such as in mines and chemical industry, variable frequency induction motor drives are natural choice. They have several other applications such as traction, mill run out tables, steel mills, fans, pumps, blowers, compressors, spindle drives, conveyers, machine tools, and so on.

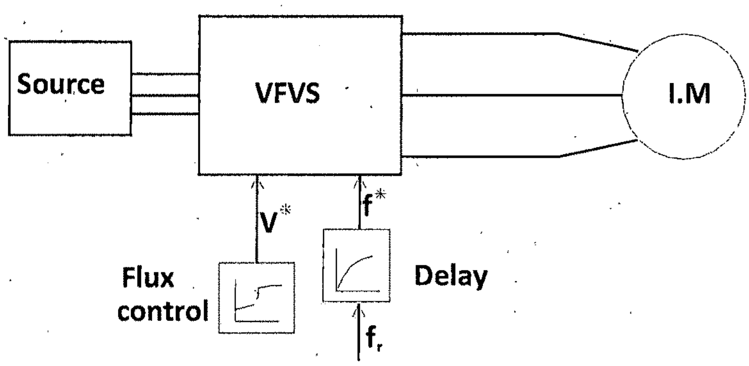


Fig.3.10 Block diagram for variable frequency control

Block diagram of variable frequency speed control scheme is shown in fig. 3.10. The motor is fed from a variable frequency voltage source (VFVS). V^* and f^* are voltage and frequency commands for VFVS. Flux control box produces a voltage command V^* for VFVS in order to maintain the relationship of fig. 3.9 between V^* and f^*. Reference frequency fr is changed to control the speed. A delay circuit is introduce between fr and f^*, so that even when fr is changed by large amount, f^* will change only slowly so that motor speed can track changes in f^*, thus restricting the motor operation for each frequency between synchronous speed and the maximum torque point. VFVS can be a voltage source inverter or a Cycloconverter.

VFDs eliminate the need for mechanical or hydraulic drives (clutches, gears, pulley, valves, and vanes). The term adjustable speed drive (ASD) is often used interchangeably with VFDs, though this is not quite accurate. ASDs include VFDs, but also methods to control the speed of dc motors, technology that has been around since the early 1960s. During the 1980s, VFDs becomes the preferred method of driving many types of load at continuously variable speeds. VFDs are used with ordinary types of ac induction motors or synchronous motors.

Working of variable frequency drives:-

The three main parts of a VFD are as follows:

1. Regulator-controls the rectifier and inverter to produce the desired ac frequency and voltage.
2. Rectifier-converts the fixed frequency ac voltage to dc.
3. Inverter-switches the rectified dc voltage to ac, creating variable ac frequency (and controlling current flow, if desired).

VFDs are totally electronic devices that convert alternating current to direct current by a rectifier. The direct current is converted back to alternating current by an inverter, at a frequency that will drive the motor at the desired speed. VDFs change the speed of the motor by using high-power semiconductor switching devices (silicon controlled rectifiers, thyristors, or power transistors). These devices change the frequency of the current that is supplied to the motor, but do so only in an on/off manner. As a result, the alternating current that is provided to the motor is not a smooth sine wave. Although this makes the motor itself less efficient, the ability of the motor to respond to load variations increases the overall energy efficiency. Without a VFD a motor might run efficiently at a single speed, but it may likely be the wrong speed as the motor's load varies.

3.4 CONTROL TECHNIQUES

Various speed control techniques implemented by modern age VFD are mainly classified in the following three categories:
[A] Scalar Control (V/f Control)
[B] Vector Control (Indirect Torque Control)
[C] Direct Torque Control (DTC)

3.5 SCALAR CONTROL

In this type of control, the motor is fed with variable frequency signals generated by the PWM control from an inverter. Here, the V/F ratio is maintained constant in order to get constant torque over entire operating range. Since only magnitudes of the input variables- frequency and voltage-are controlled, this is known as scalar control. Generally, drives with such a control are without any feedback devices (open loop control). Hence, a control of this type offers low cost and an easy to implement solution.

In such controls, very little knowledge of the motor is required for frequency control. Thus, this control is widely used. A disadvantage of such a control is that the torque developed is load dependent as it is not controlled directly. Also, the transient response of such a control is not fast due to the predefined switching pattern of the inverter. However, if there is a continuous block to the rotor rotation, it will lead to heating of the motor regardless of implementation of the over current control loop. By adding a speed/position sensor, the problem relating to the blocked rotor and the load dependent speed can be overcome. However, this will add to the system cost, size and complexity. There are a number of ways to implement scalar control. The popular schemes are described in the following sections.

3.6 SIX-STEP PWM

The inverter of the VFD has six distinct switching states. When it is switched in a specific order, the three- phase AC induction motor can be rotated. The advantage of this method is that there is no intermediate calculation required and thus, is easiest to implement. Also, the magnitude of the fundamental voltage is more than the DC bus. The disadvantage is higher low-order harmonics which cannot be filtered by the motor inductance. This means higher losses in the motor, higher torque ripple and jerky operation at low speed.

3.7 VOLTAGE SOURCE INVERTER (VSI) CONTROL

The inverter used for the control of as drives can be a voltage source inverter or a current source inverter. An inverter circuit can be a voltage source inverter if, viewed from the load side, the ac terminals of the inverter function as a voltage source. Similarly, a current source inverter is that inverter function as a current source.

Voltage source inverter has low internal impedance and thus its terminal voltage remains substantially constant with variation in load. Therefore VSI is suitable for multi-motor drives, although it is equally good for a single motor drive.

The current source inverter (CSI) because of its large internal impedance is not suitable for multi-motor drives, because any change in load on one of the motor will affect the voltage supplied to other motors also.

Since the inverter current of CSI is independent of the load impedance, it has inherent protection against short circuit across its terminals. Any short circuit in VSI may give rise to high currents and therefore a fast over current protection device is essential.

Variable frequency and variable voltage supply for induction motor control can be obtained either from voltage source inverter (VSI) or a cycloconverter.

Voltage source inverter allows a variable frequency to be obtained from a dc supply. Fig. 3.11 shows a VSI employing thyristors. Any other slef-commutated device can be used instead of transistor..

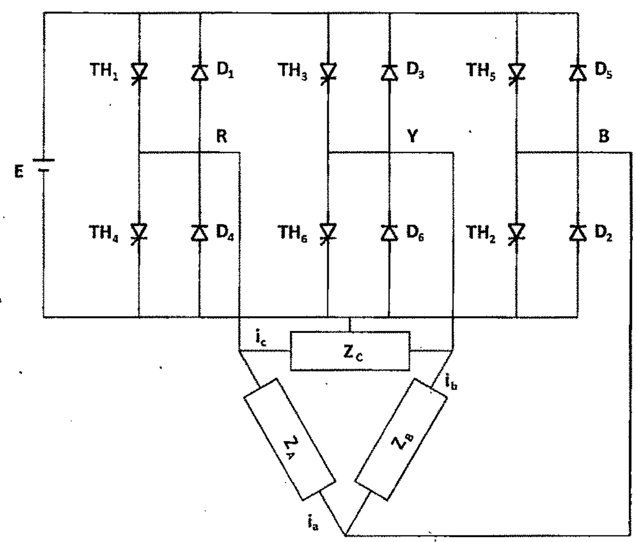


Fig. 3.11 Three Phase Bridge Inverter

Generally MOSFET is used in low voltage and low power inverter, IGBT (Insulated gate bipolar transistor) and power transistors are used up to medium power levels and GTO (gate turn off thyristor) and IGCT (Insulated gate commutated thyristor) are used for high power levels.

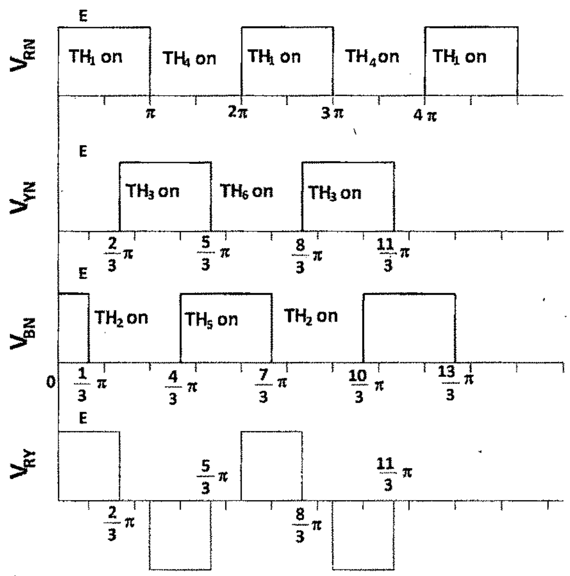


Fig. 3.12 Voltage Waveforms of the Inverter shown in fig. 3.11

VSI can be operated as a stepped wave inverter or a pulse-width modulated (PWM) inverter. When operated as a stepped wave inverter, thyristors are switched in the sequence of their numbers with a time difference of ***T/6*** and each transistor is kept on for the duration ***T/2***, where ***T*** is the time period for one cycle. Resultant line voltage wave form is shown in fig. 3.12. Frequency of inverter operation is varied by varying ***T*** and the output voltage of the inverter is varied by varying dc input voltage.

Inverter output phase voltage and line voltage is given by following Fourier series.

$$V_{RN} = \frac{2}{\pi}E\left[\sin\omega t + \frac{1}{5}\sin 5\omega t + \frac{1}{7}\sin 7\omega t\right] \_\_(3.19)$$

The rms value of the fundamental phase voltage is given by

$$V_P = \frac{\sqrt{2}}{\pi}E \_\_\_\_\_(3.20)$$

$$V_{RY} = \frac{2\sqrt{3}}{\pi}E\left[\sin\omega t - \frac{1}{5}\sin 5\omega t - \frac{1}{7}\sin 7\omega t + \frac{1}{11}\sin 11\omega t + \frac{1}{13}\sin 13\omega t \ldots\ldots\right] \_\_(3.21)$$

The rms value of the fundamental line voltage is given by

The rms value of the line voltage given by eq. (3.21) is

$$V_{TL} = \sqrt{\frac{2}{3}}E = 0.816E\_\_(3.22)$$

Therefore, the total harmonic component is not more than 31.08 percent of the fundamental. By using more complex circuits having more number of thyristors, the number of steps in the voltage waveform can be increased which makes the wave forms more ideal.

If the load is star connected instead of delta, the voltage waveform with respect to neutral point is as shown in fig. 3.13. The Fourier analysis of this waveform yields the following expression:

$$V_{RO} = \frac{3}{\pi}E\left[\sin\omega t - \frac{1}{5}\sin 5\omega t - \frac{1}{7}\sin 7\omega t + \frac{1}{11}\sin 11\omega t + \frac{1}{13}\sin 13\omega t \ldots\ldots\right]\_\_(3.23)$$

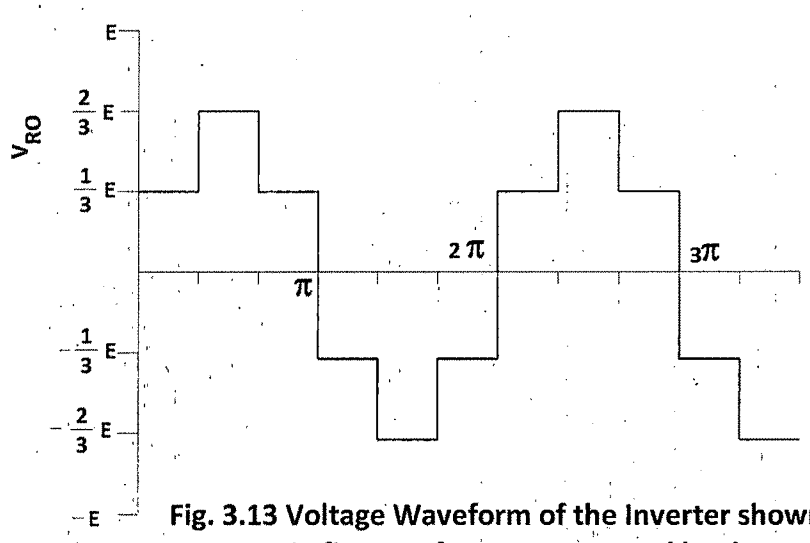


Fig. 3.13 Voltage Waveform of the Inverter shown in fig. 3.11 for star connected load

The current waveforms consist of series of exponentially rising or decreasing components depending on the switching of the thyristors.

3.8 VOLTAGE CONTROL OF INVERTERS

To maintain a constant flux density in the induction motor, the inverter must maintain constant ration of voltage to frequency. This voltage variation can be obtained by one of the following possible methods:

[1] By varying direct voltage input of the inverter.

[2] By varying output voltage of the inverter.

[3] By using switching techniques within the inverter.

In a dc link inverter, the magnitude of its output ac voltage depends on the magnitude of its input dc voltage. Therefore by having a variable dc supply obtained from the converter, the output voltage of the inverter can be varied.

When supply is dc, variable dc input voltage is obtained by connecting a chopper between dc supply and inverter (fig.3.14(a)). When supply is ac, variable dc input voltage is obtained by connecting a controlled rectifier between ac supply and inverter (3.14(b)). A Large electrolytic filter capacitor ***C*** is connected in dc link to make inverter operation independent of rectifier or chopper and to filter out harmonic in dc link voltage.

For controlling the output voltage of the inverter from its outside, a variable ratio output transformer may be used. The output of the inverter is connected to input terminals of the transformer. The voltages tapping of the transformer are controlled automatically using closed loop control.

The third method of voltage control requires control of switching in the thyristors to modify the output wave form. The most common method used for this type of control is by having pulse width modulated inverter.

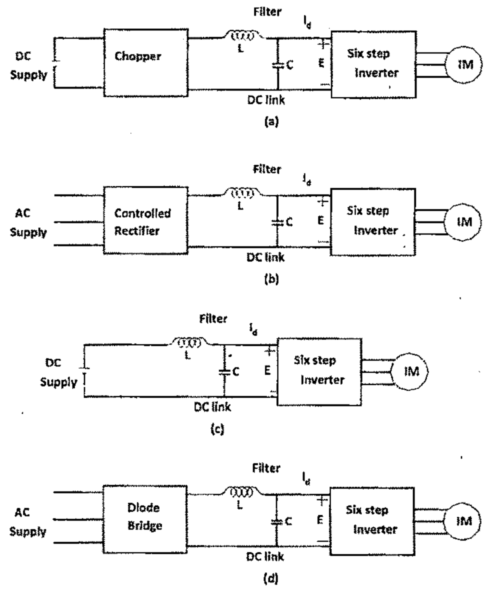


Fig. 3.14 VSI controlled IM drives

The main drawback of steeped wave inverter is the large harmonics of low frequency in the output voltage. Consequently, an induction motor drive fed from a stepped wave inverter suffers from the following drawbacks:

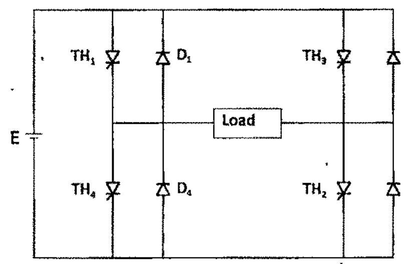
[1] Because of low frequency harmonics, the motor losses are increased at all speeds causing derating of the motor.

[2] Motor develops pulsating torques due to fifth, seventh, eleventh and thirteenth harmonics which cause jerky motion of the rotor at low speeds.

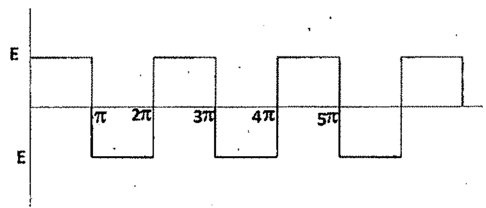
[3] Harmonic content in motor current increases at low speeds. The machine saturates at light loads at low speeds due to high *(V/f)* ratio. These two effects overheat the machine at low speeds, thus limiting lowest speed to around 40% of the base speed.

Harmonics are reduced, low frequency harmonics are eliminated, associated losses are reduced and smooth motion is obtained at low speed also when inverter is operated as a pulse-width modulated inverter.

The method of voltage control using the pulse width modulation scheme can be explain with the help of single phase bridge inverter of fig. 3.15. The output voltage waveform of this inverter is shown in fig. 3.15(b). It is a square wave having +E magnitude for $0 < \omega t < \pi$ and −E for $\pi < \omega t < 2\pi$. If the voltage control is desired the output pulse may be controlled as shown in fig 3.15(c).



(a) Single phase bridge inverter



(b) Output voltage waveform of a single phase bridge inverter

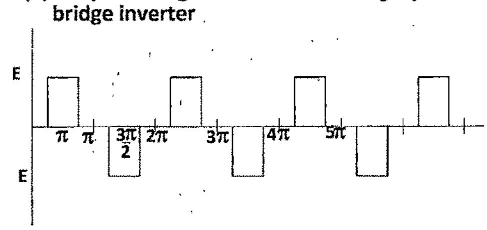


Fig. 3.15 Single phase bridge inverter and its output voltage waveforms

The thyristors connecting the load to the supply are not always 'ON' but there is some period for which no supply is connected to the load. In a single pulse modulation, a pulse width is created around $\pi/2$ and $3\pi/2$; the variation in voltage is obtained by varying the width of these pulses. As can be seen from fig.3.15, the harmonic contents are increases for low voltage output of the inverter. Therefore, multiple pulse modulation schemes are preferred. In a multiple pulse scheme, instead of one pulse in each half cycle, there are several equidistant pulses as shown in fig. 7.16 are generated. As shown in figure, there are ***M*** pulses, each of height ***E***, and the variation in voltage magnitude is obtained by varying the pulse width. This type of modulation is achieved by comparing a dc level with triangular wave. By varying the dc voltage level, the pulse width can be varied.

The advantage of multi pulse modulation is that the voltage control is achieved with simultaneous reduction of lower order harmonics. However at very low frequency, and for constant V/F the out voltage wave form deteriorates, since the interval between pulses increases.

3.9 SINUSOIDAL PWM

weighted values are used. Here the pulse width is made the sinusoidal function of the angular position of the pulse in the cycle. The advantage of this technique is that very little calculation is required. Only on look-up table of sine wave is required, as all the motor phases are 120^0 (electrical) displaced. The disadvantage of this method is that the magnitude of the fundamental voltage is less than 90%. Also, the harmonics at PWM switching frequency have significant magnitude

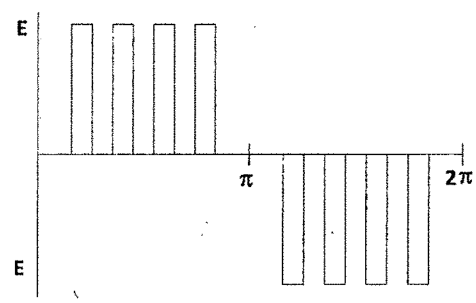


Fig. 3.16 Multipulse modulation scheme

Thus the pulse width increases as the angle increases from ***0*** to ***90^0*** then the width decreases from ***90^0*** to ***180^0*** again the pulse width increases in negative direction from ***180^0*** to ***270^0*** then the width decreases from ***270^0*** to ***360^0*** . The output voltage waveform of sinusoidal pulse width modulated inverter is shown in fig.3.17. The waveforms can be generated by means of a control circuit in which a high frequency triangular wave form is compared with a rectified sinusoidal wave.

Since output voltage can now be controlled by pulse width modulation, no arrangement is required for the variation of input dc voltage, hence inverter can be directly connected when the supply is dc [fig. 3.14(c)] and through a diode rectifier when supply is ac [fig. 3.14(d)].

The fundamental component in the output phase voltage of a PWM inverter operating with sinusoidal PWM is given by

$$V = m\frac{E}{2\sqrt{2}} \quad \text{\_\_\_\_\_\_\_\_(3.24)}$$

Where ***m*** is the modulation index.

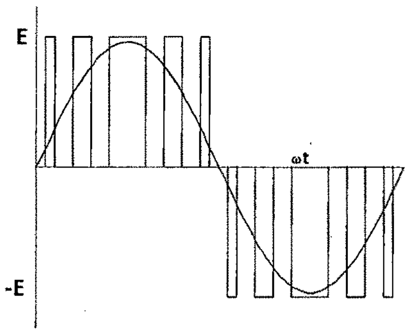


Fig. 3.17 Sinusoidal Pulse width modulated inverter

The harmonics in the motor current produce torque pulsation and derate the motor. For a given harmonic content in motor terminal voltage, the current harmonics are reduced when the motor has higher leakage inductance; this reduces derating and torque pulsations. Therefore, when fed from VSI, induction motors with large (compared to when fed from sinusoidal supply) leakage inductance is used.

3.10 CYCLOCONVERTER

In cycloconverter the ac voltage at supply frequency is directly converted to a voltage source of lower frequency. Normally cycloconverters are used to get a single phase or a three phase supply output from a three phase input. Basically it consists of two group of phase controlled rectifier circuit for each phase of the output, one group for the positive portion and the other group for the negative portion. By delaying the firing angle of the controlled rectifier the mean output voltage of the rectifier can be controlled. If the rectifier firing angle is slowly varied from ***90*** degree to ***0*** degree and ***0*** degree to ***90*** degree, the mean output voltage will vary from zero to maximum value, then from maximum to zero. Therefore, it is possible to have low frequency sinusoidal variations superimposed on the output voltage of the rectifier. A second group of the rectifiers is used to produce negative half cycle of the waveform. The frequency of the superimposed voltage is totally dependent on the firing angle control and is independent of the supply frequency. A three-phase to single phase cycloconverter is shown in fig. 3.18.

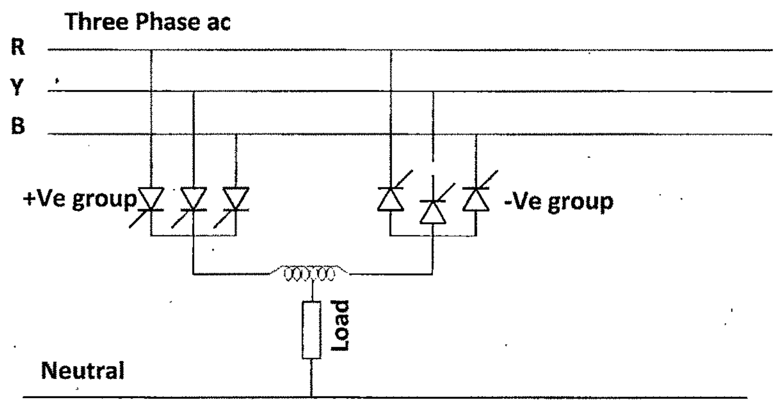


Fig 3.18 Three Phase to Single Phase cycloconverter

The variation of firing angle for the positive group rectifier and for the negative group rectifier is shown in fig. 3.19(a) and3.19 (b) respectively. As can be seen from fig. 3.19(a), the firing angle at ***H*** is $\pi/2$ and average value of output voltage is zero. The firing angle is decreased at ***J, K,*** and it is zero at ***L***. The firing angle is again increased at ***M, N, O*** and it $\pi/2$ at P. The average value of the voltage is shown as broken line. Similarly, for the negative group of the rectifier bridge, by controlling the firing angle from 0 to $\pi/2$, the mean output voltage can be obtained which has lower frequency as shown in fig. 3.19(b).

The average back emf of the inverter may also be controlled in a sinusoidal form by varying the firing angle from $\pi/2$ **to** π and the power flow can be reversed. Thus it is possible to have regenerative operation of the induction motor fed from a cycloconverter.

Since the positive and negative group of rectifiers are connected in inverse parallel, their average output voltage must be equal. This is achieved by making the firing angle for the negative group $\propto_n = \pi - \propto_p$ where $\propto_p$ is firing angle for the positive group. However, instantaneous output voltages of the two groups are not equal and a centre-tapped reactor is used to limit the circulating current.

For three phases to three cycloconverter circuits, three single phase cycloconverters with a phase displacement of 120^0 between their outputs are used as shown in fig. 3.20. The cycloconverter has certain advantages as well as disadvantages over dc link inverter. It has the following advantages.

[1] the cycloconverter is more efficient as there is only one stage of conversion compared to two stages of conversion required in dc link inverters. Also the cycloconverter does not require forced commutation and nearly 90 % efficiency can be obtained.

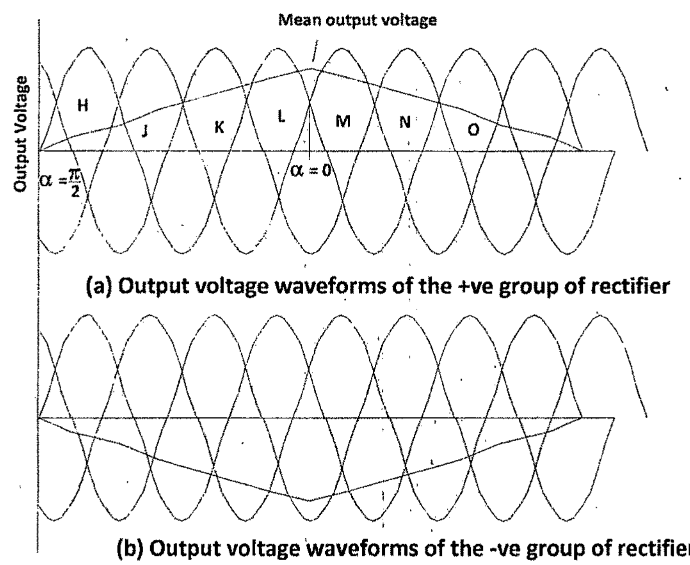


Fig 3.19 Output voltage waveforms of cycloconverter

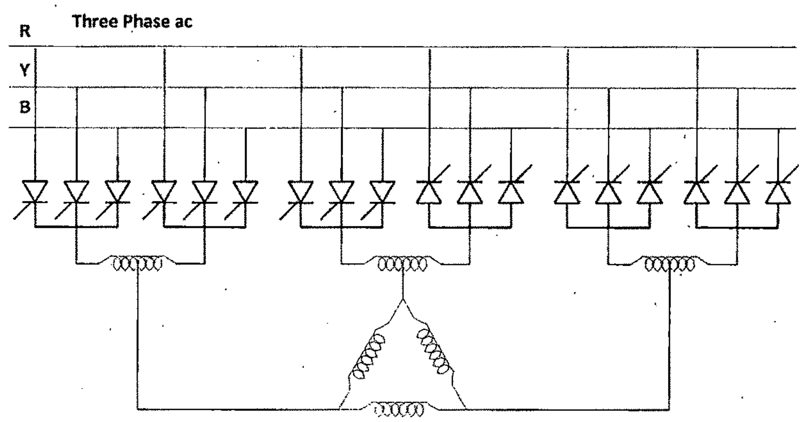


Fig 3.20 Three Phase to Three Phase cycloconverter

[2] The output waveform of cycloconverter voltage is almost pure sinusoidal whereas most dc link inverters produce stepped voltage waveform.

[3] The cycloconverter is capable of power transfer in both directions, therefore regenerative braking is possible.

The main disadvantage of the cycloconverter is that the maximum output frequency must be less than one third of the input frequency and, therefore, the minimum speed of the drive motor is limited to 1/3 of the speed corresponds to input frequency. The cycloconverter also requires more thyristors than the dc link inverter and, therefore, is not economical for low power motors.

Thus cycloconverter is more useful for high power low speed reversible drives whereas the dc link inverter is more suitable for high speed operations.

3.11 CONSTANT VOLTS/HERTZ INDUCTION MOTOR DRIVES

The operation of induction machines in a constant volt per hertz mode back to the late fifties and early sixties but were limited in their low speed range [21]. Today constant volt per hertz drives are built using PWM-IGBT-based inverters of the types discussed earlier and the speed range has widened to include very low speeds [22] although operation very near zero speed (less than 1 Hz) remains as a challenge mainly due to inverter non-linearties at low output voltages.

Ideally, by keeping a constant V/f ratio for all frequencies the nominal torque-speed curve of the induction motor can be reproduced at any frequency as discussed in Section 3.2. Specifically if stator resistance is neglected and keeping a constant slip frequency the steady state behaviour of the induction machine can be characterized as impedance proportional to frequency. Therefore, if the V/f ratio is kept constant the stator flux, stator current, and torque will be constant at any frequency. This feature suggests that to control the torque one needs to simply apply the correct amount of V/Hz to stator windings. This simple, straight forward approach, however, does not work well in reality due to several factors, the most important ones being

1) Effect of supply voltage variations
2) Influence of stator resistance
3) Non-ideal torque/speed characteristic (effects of slip)
4) Non-linearties introduced by the PWM inverter.

Low frequency operation is the particularly difficult to achieve since these effects are most important at low voltages. Also, the non-linearties within the inverter, if not adequately compensated, yield highly distorted output voltages which, in turn, produces pulsating torques that lead to vibrations and increased acoustic noise.

In addition to these considerations, a general purpose inverter must accommodate a variety of motors from different manufacturers. Hence it must compensate for the above mentioned effects regardless of machine parameters. The control strategy must also be capable of handling parameter variations due to temperature and/or saturation effects. This fact indicates that in a true general purpose inverter it is necessary to include some means to estimate and/or measure some of the machine parameters. Another aspect that must be considered in any practical implementation deals with the DC bus voltage regulation, which, if not taken into account, may lead to large errors in the output voltage.

Because general purpose drives are cost sensitive it is also desired to reduce the number of sensing devices within the inverter. Generally speaking only the dc link inverter voltage and current are measured; hence the stator current and voltage must be estimated based only on these measurements. Speed encoders or tachometers are not used because they add cost as well as reduce system reliability.

Other aspects that must be considered in the implementation of an ideal constant *V/f* drive relate to:

a) Current measurement and regulation,
b) Changes in gain due to pulse dropping in the PWM inverter,
c) Instabilities due to poor volt-second compensation that result in lower damping. This problem is more important in high efficiency motors, and
d) Quantization effects in the measured variables.

Another aspect that must be carefully taken into account is the quantization effect introduced by the A/D converters used for signal acquisition. A good cost to resolution compromise seems to be the use of 10 bit converters. However, a high performance drive is likely to require 12 bit accuracy.

3.12 COMPENSATION FOR SUPPLY VOLTAGE VARIATIONS

In an industrial environment, a motor drive is frequently subjected to supply voltage fluctuations which, in turn, imposed voltage fluctuations on the DC link of the inverter. If these variations are not compensated for, the motor will be impressed with either and under or an overvoltage which produces excessive I^2R loss or excessive iron loss respectively. The problem can be avoided if the DC link voltage is measured and the voltage command V^* adjusted to produce a modified command V^{**} such that

$$V^* = \left(\frac{V_{busR}}{V_{bus}}\right) V^{**} \_\_\_\_\_\_ (3.25)$$

$$Where\ V_{busR} = Rated\ value\ of\ bus\ voltage.$$

3.13 IR COMPENSATION

A simple means to compensate for the resistive drop is to boost the stator voltage by I_1R_1 (voltage proportional to the current magnitude) and neglect the effect of the current phase angle. To avoid the direct measurement of the stator current this quantity can be estimated from the magnitude of the dc-link current [23]. In this paper a good ac current estimate was demonstrated at frequencies as low as 2 Hz but the system requires high accuracy in the dc-link current measurement making it impractical for low cost applications. A robust ***IR*** boost method must include both magnitude and phase angle compensation. Typically currents of two phases must be measured with the third current inferred since the currents sum to zero. In either case the value of the stator resistance must be known.

The value of the stator resistance can be estimated by using any one of several known techniques [24]-[26]. Unfortunately these parameter estimation techniques require knowing the rotor position or velocity and the stator current. An alternate method of `boosting' the stator voltage at low frequencies is presented in [27]. Here the *V/f* ratio is adjusted by using the change in the sine of the phase angle of motor impedance. This approach also requires knowing the rotor speed and it is also dependent on the variation of the other machine parameters. Its practical usefulness is questionable because of the technical difficulty of measuring phase angles at frequencies below 2 Hz.

Constant Volts/Hz control strategy is typically based on keeping the stator flux linkage magnitude constant and equal to its rated value. Using the steady state equivalent circuit of the induction motor, shown in Figure 3.5, an expression for stator voltage compensation for resistive drop can be shown to be

$$V_1 = \frac{\sqrt{2}}{3} I_{1Re}\widehat{R_1} + \sqrt{\frac{V_{1R}f_e}{f_R} + \frac{2}{9}\left(I_{1Re}\widehat{R_1}\right)^2 - \left(I_1\widehat{R_1}\right)^2} \_\_\_\_\_\_ (3.26)$$

Where V_{1R} is the base (rated) rms phase voltage at base frequency, f_R and f_e are the rated and line frequency in Hertz, $\widehat{R_1}$ is the estimated value of resistance, I_s is the rms current obtained on a instantaneous basis by,

$$I_s = \sqrt{\frac{2}{3}}\sqrt{i_a(i_a + i_c) + i_c^2} \qquad (3.27)$$

And I_{1Re} is the real component of rms stator current obtained from

$$I_{1Re} = i_a\left[\cos\theta_e - \cos\left(\theta_e - \frac{2\pi}{3}\right)\right] + i_c\left[\cos\left(\theta_e + \frac{2\pi}{3}\right) - \cos\left(\theta_e - \frac{2\pi}{3}\right)\right] \qquad (3.28)$$

where i_a *and* i_c are two of the instantaneous three phase stator currents, $\theta_e = \omega_e t$ and the cosine terms are obtained from the voltage command signals. The estimated value of resistance can be obtained either by a simple dc current measurement corrected for temperature rise or by a variety of known methods [24]-[26]. Derivation details of these equations are found in [28]. Given the inherently positive feedback characteristic of an *Ir* boost algorithm it is necessary to stabilize the system by introducing a first order lag in the feedback loop (low-pass filter).

3.14 SLIP COMPENSATION

By its nature, the induction motor develops its torque as a rotor speed slightly lower than synchronous speed (effects of slip). In order to achieve a desired speed, the applied frequency must therefore be increased by an amount equal to the slip frequency. The usual method of correction is to assume a linear relationship exists between torque and speed in the range of interest, Hence, the slip can be compensated by knowing this relationship. This approximation gives good results as long as the breakdown torque is not approached. However, for high loads the relationship becomes non-linear. Ref. [28] describes a correction which can be used for high slip,

$$f_{slip} = \frac{1}{2 - A * P_{gap}}\left\{\sqrt{f_e^{*2} + \frac{\frac{S_m}{S_R} S_{linear}}{2 * \frac{T_{bd}}{T_R}} P_{gap} - B * P_{gap}^2} - f_e^*\right\} \qquad (3.29)$$

where f_e^* is the external command frequency and,

$$A = \frac{P}{4\pi S_{bd} T_{bd} f_R} \qquad (3.30)$$

and

$$B = \left(\frac{P}{4\pi T_{bd}}\right)^2 \text{\_\_\_\_\_\_}(3.31)$$

and P is the number of poles. The slope of linear portion of the torque speed curve is given by

$$s_{linear} = \frac{P s_R f_R}{\pi T_R} \text{\_\_\_\_\_\_}(3.32)$$

Finally the air gap power is

$$P_{gap} = 3V_1 I_1 (pf) - 3I_1^2 \widehat{R_1} - P_{core} \text{\_\_\_\_\_\_}(3.33)$$

where P_{core} at rated frequency can be obtained from

$$P_{coreR} = P_{inR}\left(1 - \frac{\eta_R}{1 - S_R}\right) - 3I_1^2 \widehat{R_1} \text{\_\_\_\_\_\_}(3.34)$$

Where the quantities $S_R, f_R, \eta_R, I_1, P_{inR}$ and T_R are the rated values of slip, line frequency, efficiency, stator current, input power and torque respectively. All of these quantities can be inferred from the name plate data.

3.15 VOLT-SECOND COMPENSATION

One of the main problems in open-loop controlled PWM-VSI drives is the non-linearity caused by the non-ideal characteristics of the power switches. The most important non-linearity is introduced by the necessary blanking time to avoid short circuiting the DC link during the commutations. To guarantee that both switches are never on simultaneously a small time delay is added to the gate signal of the turning-on device. This delay, added to the device's inherent turn-on and turn-off delay times, introduces a magnitude and phase error in the output voltage [29]. Since the delay is added in every PWM carrier cycle the magnitude of the error grows in proportion to the switching frequency, introducing large errors when the switching frequency is high and the total output voltage is small.

The second main non-linear effect is due to the finite voltage drop across the switch during the on-state [30]. This introduces an additional error in the magnitude of the output voltage, although somewhat smaller, which needs to be compensated.

To compensate for the dead-time in the inverter it is necessary to know the direction of the current and then change the reference voltage by adding or subtracting the required volt-seconds. Although in principle this is simple, the dead time also depends on the magnitude and phase of the current and the type of device used in the inverter. The dead-time introduced by the inverter causes serious waveform distortion and fundamental voltage drop when the switching frequency is high compared to the fundamental output frequency. Several papers have been written on techniques to compensate for the dead time [29],[31]-[33].

Regardless of the method used, all dead time compensation techniques are based on the polarity of the current, hence current detection becomes an important issue. This is specially true around the zero-crossings where an accurate measurement is needed to correctly compensate for the dead time. Current detection becomes more difficult due to the PWM noise and because the use of filters introduces phase delays that needed to be taken into account.

The name "dead-time compensation" often misleads since the actual dead time, which is intentionally introduced, is only one of the elements accounting for the error in the output voltage, for this reason here it is referred as volt-second compensation.

3.16 SPACE VECTOR MODULATION PWM (SVMPWM)

This control technique is based on the fact that three-phase voltage vectors of the induction motor can be converted into a single rotating vector. Rotation of this space vector can be implemented by VFD to generate three-phase sine waves. The advantages are less harmonic magnitude at the PWM switching frequency due to averaging, less memory requirement compared to sinusoidal PWM, etc. The disadvantages are not full utilization of the DC bus voltage, more calculation required, etc.

3.17 SVMPWM WITH OVERMODULATION

Implementation of SVMPWM with over modulation can generate a fundamental sine wave of amplitude greater than the DC bus level. The disadvantage is complicated calculation, line-to-line waveforms are not clean and the THD increases, but still less than the THD of the six-step PWM method.

3.18 VECTOR CONTROL

The various techniques for producing a variable frequency supply source for controlling speed-torque characteristic for an induction motor have provided good steady state response. The constant V/F control technique provides constant flux and hence constant torque for the entire range of operation. However at low speed i.e. at low frequencies the steady state response is not as good as at higher frequencies.

The dynamic response of the drive is poor. This poor dynamic response is because of the deviation of air gap flux linkage from their set values. This deviation is not only in magnitude but also in phase. The variation in the flux linkages have to control by the magnitude and frequency of the stator and rotor phase current and their instantaneous phases. So far, the control techniques have utilized the stator phase current magnitude and frequency and not their phases. This resulted in the deviation of the phase and magnitudes of the air gap flux linkages from their set values.

The oscillations in the air gap flux linkages result in oscillations in electromagnetic torque oscillations. If phenomenon is neglected (i.e. not attended) then it results in speed oscillations. This is undesirable in many high-performance applications, such as in robotic actuators, centrifuges, servos, process drives and metal rolling mills, where high precision, fast positioning, or speed control are required. Such requirement will not be met with the sluggishness of control due to the flux oscillations. Further, air gap flux variations result in large variation of stator currents, requiring large peak converter and inverter ratings to meet the dynamics. An enhancement of peak inverter rating increases cost and reduces the competitive edge of ac drives in the marketplace, in spite of excellent advantages of the ac drives over dc drives.

Vector control is also known as the field oriented control or flux oriented control or indirect torque control. Using field orientation (Clarke-Park transformation), three phase current vectors are converted to a two dimensional rotating reference frame (d-q) from a three-dimensional stationary reference frame. The ***d*** component represents the flux producing component of the stator current and the ***q*** component represents the torque producing component. These two decoupled components can be independently controlled by passing though separate ***PI*** controllers. The outputs of the ***PI*** controllers are transformed back to the three-dimensional stationary reference plane using the inverse of the Clarke-Park transformation. The corresponding switching pattern is pulse width modulated and implemented using the SVM.

This control simulates a separately exited DC motor model, which provides an excellent torque-speed curve. The transformation from the stationary reference frame to the rotating reference frame is done and controlled with reference to a specific flux linkage space vector (stator flux linkage, rotor flux linkage or magnetizing flux linkage). In general, there exists three possibilities for such selection and hence, three different vector controls. They are:

[1] Stator flux oriented control

[2] Rotor flux oriented control

[3] Magnetizing flux oriented control

As the torque producing component in this type of control is controlled only after transformation is done and is not the main input reference, such control is known as indirect torque control. The most challenging and ultimately, the limiting feature of the field orientation, is the method whereby the flux angle is measured or estimated. If the field angle is calculated by using terminal voltages and currents or Hall sensors or flux sensing windings, then it is known as ***direct vector control***. The field angle can also be obtained by using rotor position measurement and partial estimation with only machine parameters but not any other variables, such as voltages or currents; using this field angle leads to a class of control schemes known as ***indirect vector control***.

In *direct vector control*, the flux measurement is done by using the flux sensing coils or the Hall devices. This adds to additional hardware cost and in addition, measurement is not highly accurate. Therefore, this method is not a very good control technique. The more common method is *indirect vector control*. In this method, the flux angle is not measured directly, but is estimated from the equivalent circuit model and from measurements of the rotor speed, the stator current and the voltage.

One common technique for estimating the rotor flux is based on the slip relation. This requires the measurement of the rotor position and the stator current. With current and position sensors, this method performs reasonably well over the entire speed range. The most high-performance VFDs in operation today employ indirect field orientation based on the slip relation. The main disadvantage of this method is the need of the rotor position information using the shaft mounted encoder. This means additional wiring and component cost. This increases the size of the motor. When the drive and the motor are far apart, the additional wiring poses a challenge.

To overcome the sensor/encoder problem, today's main research focus is in the area of a sensor less approach. The advantages of the vector control are better torque response compared to the scalar control, full-load torque close to zero speed, accurate speed control and performance approaching DC drive, among others. But this requires a complex algorithm for speed calculation in real-time. Due to feedback devices, this control becomes costly compared to the scalar control.

3.19 DIRECT TORQUE CONTROL (DTC)

The difference between the traditional vector control and the DTC is that the DTC has no fixed switching pattern. The DTC switches the inverter according to the load needs. Due to elimination of the fixed switching pattern (characteristic of the vector and the scalar control), the DTC response is extremely fast during the instant load changes. Although the speed accuracy up to 0.5% is ensured with this complex technology, it eliminates the requirement of any feedback device. The block diagram of the DTC implementation is shown in Figure 24. The heart of this technology is its adaptive motor model. This model is based on the mathematical expressions of basic motor theory. This model requires information about the various motor parameters, like stator resistance, mutual inductance, saturation coefficient, etc. The algorithm captures all these details at the start from the motor without rotating the motor. But rotating the motor for a few seconds helps in the tuning of the model. The better tuning gives higher accuracy of speed and torque control. With the DC bus voltage, the line currents and the present switch position as inputs, the model calculates actual flux and torque of the motor. These values are fed to two-level comparators of the torque and flux, respectively. The output of these comparators is the torque and flux reference signals for the optimal switch selection table. Selected switch position is given to the inverter without any modulation, which means faster response time. The external speed set reference signal is decoded to generate the torque and flux reference. Thus, in the DTC, the motor torque and flux become direct controlled variables and hence, the name Direct Torque Control.

The advantage of this technology is the fastest response time, elimination of feedback devices, reduced mechanical failure, performance nearly the same as the DC machine without feedback, etc. The disadvantage is due to the inherent hysteresis of the comparator, higher torque and flux ripple exist. Since switching is not done at a very high frequency, the low-order harmonics increases. It is believed that the DTC can be implemented using an Artificial Intelligence model instead of the model based on mathematical equations. This will help in better tuning of the model and less dependence on the motor parameters.