CHAPTER 1

INTRODUCTION

1.1 HISTORY AND OVERVIEW

Over the past decades DC machines were used on large scale for variable speed applications due to the decoupled control of torque and flux which can be obtained by armature and field current control respectively. DC drives are useful in many aspects as high starting torque and ease of control. In spite of all these advantages DC motor drives suffer from one major drawback which is the presence of mechanical commutator and brush.

The robustness, low cost, enhanced performance and ease of maintenance make the motor drives more advantageous in so many applications [1-3]. The DTC principle was introduced in early days of second half of 1980s. In 1986, Takahashi and Noguchi developed a new methodology to control torque and flux using PWM output voltage which provides faster response and is highly efficient. The concept of DTC spread widely and quickly than vector control in industry applications. The most frequently used static power converter in DTC drive is VSI.

The control mechanism of DTC having two parallel branches like vector control. Flux and torque set values are two references which may or may not be the output of speed controller, depending on control methodology of drive either torque or speed controlled. Estimation of stator flux and motor torque is required for close loop control of torque and flux. The errors between the estimated and set values are used in totally different way as compared with vector control [13-14]. The main idea behind DTC is that the error in torque and flux is directly used to operate the inverter without any intermediate current control loop and co-ordinate transformation. Hysteresis controller output is used to obtain the possible inverter switching modes to be applied to inverter terminals such that the errors in flux and torque should lie in the prescribed limits [26].

A direct torque controlled induction motor drive was first commercially manufactured by ABB in mid 1990s. The core of the control system in DTC is the subsystem with torque and hysteresis controller and inverter switching logic. An exact model of machine is also required because estimation of stator flux and motor torque is based on machine model and machine input stator current and voltages and stationary frame reference is used to build DTC theory.

When resistance of stator winding is neglected then stator flux is integral of the applied voltage. Hence in short duration the change in stator flux is proportional to applied voltage. The time constant of rotor of induction motor is normally large hence the rotor flux linkage will change slowly as compared to the stator flux linkage. The forward active voltage space vector is applied when the required torque is to be increased and this causes the torque angle increase. Similarly backward or zero active voltage space vector is applied when the torque requirement is less than as present and this causes the torque angle to be decreased. Hence it can be concluded that the torque can be controlled directly and increased and decreased directly by rotating the stator flux linkage space vector to the appropriate position. This is why the control scheme is called DTC [5].

The main advantages of DTC as compared to the vector control are:

- Direct control of flux and torque
- Co-ordinate transformation is not required
- Current controller is not required
- DTC is only sensitive to changes in stator resistance
- It's a sensor less control

The main disadvantages of DTC are:

- Torque and flux estimators are required
- Inherent torque and flux ripples are present
- Possibility of problems in starting

1.2 LITERATURE REVIEW

In **1986, Takahashi and Noguchi** [1] proposed a new technique of induction motor control based on limit cycle control of torque and flux. He used optimized PWM output voltage producing highly efficient and fast torque response.

In 1988, Depenbrock, M. [2] proposed a new methodology of DTC of induction motor drive using signal processing. The stator current and flux linkages are processed and depending on that voltage source inverter was triggered via Schmitt trigger.

In 1991, Holtz, J. And Khambadkone, A,et.al [3,4] proposed a hardware control methodology made up of 80196 microcontroller and ASIC (application specific integrated circuit). Pulse width modulated signals are generated using the proposed method used for controlling of induction motor drives.

In **1992, Thomas G Habetler et.al,** [5-7] proposed a new methodology for DTC of induction motor drive using predictive and deadbeat control of torque and flux. The variation in torque and flux is calculated by the estimation of synchronous speed and stator voltage required for matching of torque and flux values with their reference values.

In 1993, Thomas G. Habetler, Pagni et.al, [8,9], proposed a new technique of DTC of induction motor drive to improve the torque response of a large capacity induction motor using two sets of three phase inverter and an open delta induction motor. Instantaneous voltages are applied through inverter having characteristics to provide ease for selection of inverter switching mode.

In 1993, Abe T. et.al, [10-11] proposed direct torque control method based on deadbeat control of torque and stator flux using the principle of stator field flux orientation. The performance of induction motor drive was observed.

In 1995, Mir, S.A., Levi, et.al [16-18]**,** proposed a methodology of variable duty ratio control scheme to increase switching frequency, and adjust the width of hysteresis bands of conventional DTC according to the switching frequency. This technique minimizes torque and current ripples, improves torque response, and reduces switching losses instead of its simplicity.

In 1996, Tiitinen et.al [19-20] proposed ACS600, the first application utilizing DTC by using optimum switching determined by every control cycle on a time level of 25*micro*seconds.

In 1996, Blasko, V., [21] proposed a new algorithm suitable for implementation of modified space vector control on digital or analog hardware for triangle comparison PWM.

In 1997, **James N Nash** [22,23] proposed a methodology for vector control of induction motor without using encoder. The adaptive motor model was presented and reliability of control method was also discussed.

In 1998, C. Lascu, I. Boldea and F. Blaabjerg [24,25] proposed a methodology for direct torque and flux control based on space vector modulation of induction motor sensor less drives. Acoustical noise, torque, fluxes, current and speed pulsations are reduced substantially during steady state. The better steady state performance is observed in sensorless control implementation along with the DTC transient merits preserved.

In 1998, D. Casadei and G. Serra [26] proposed a new methodology of direct torque control of induction motor drive using matrix converter used for generating the voltage vector under unity power factor operation. This control method combines the advantages of matrix converters with the advantages of the DTC schemes.

In 1998, P. C. Costa, Adriano S. C., M. F. Chouzal and C. A. Martins [27] proposed a DTC controller having uniform switching frequency based on frequency domain concepts and compared its characteristics with scalar and vector control methods. This proposed method reduces the problem associated with switching of voltage source inverter.

In 1999, Jun-Koo-Kang, Dae-Woong Chug, and Seung-Ki, et.al [28-29] investigated the effects of hysteresis bands on (DTC) of an induction machine and proposed a method to control the switching frequency of inverter by variable amplitude of the hysteresis. The problem of unpredictable inverter switching frequency was compensated which varies due to operating speed, load condition and parameters of the induction machine by increasing the amplitude of hysteresis band.

In 1999, Hoang Le-Huy, et.al [30-32] compared induction motor drives field oriented control and direct torque control based on various criteria including basic control characteristics, dynamic performance, parameter sensitivity, and implementation complexity.

In 2000,Victor perelmuter, Arias, A. et.al [32-36] proposed a method to improve the dynamic performance of induction motor by rotating the stator flux vectors near the sector boundaries within prescribed hysteresis band. This also improves the stator flux in machine and harmonic contents in stator currents along with the reduced switching frequency.

In 2002, Rodic, M, et.al [41] proposed a method to direct torque control of induction motor by calculating the torque and flux derivative depending on torque, rotor flux and speed.

In 2003, Lukko, et.al, [42-43] proposed improved integration method for estimation of stator flux linkage for speed and position sensor less direct torque controlled AC machine drives, based on monitoring the scalar product of the estimated stator flux linkage and the measured stator current.

In 2004, M.vasudevan and Dr. R.Arumugan [44], proposed a new methodology of direct torque control of induction motor for electric vehicles. The comparison among field Oriented Control, Direct Torque Control (DTC), and DTC using Space Vector Modulation are done and concluded that DTC of induction motor drive using Space vector modulation is the best scheme for electric vehicle application.

In 2004, Ojo, et.al, [45] proposed switching table based hysteresis DTC, direct self-control, constant switching frequency DTC with space vector modulation.

In 2005, Srinivasan, et.al, [48] proposed software implementation for a two level inverter using space vector modulation combining MATLAB & PSIM. The switching voltage vectors were generated using MATLAB codes.

In 2005, M.Vasudevan, R.Arumugan and S.Paramasivam [46-48], proposed adaptive intelligent torque control strategy of induction motor with its advantages and disadvantages. The performance of neural network, fuzzy and genetic algorithm based torque controllers is evaluated as the various sensorless DTC techniques of IM. These adaptive intelligent techniques are applied to achieve high performance decoupled flux and torque control.

In 2006, Rahman, et.al, [49] simulated the direct torque controlled electric drive with multi-inertia mechanical model.

In 2006, Singh, B., Ravi, J., [50] proposed an optimal switching strategy for matrix converter fed induction motor drive to minimize the torque ripple and switching frequency.

In 2006, Sunter, et.al, [51,52] proposed a new simplified SVPWM method for a three level inverter having large no. of switching states compared to a two level inverter. This method can also be applied for multi-level inverter.

In 2008, Sarat K Sahoo, Tulsiram Das, Vedam Subrahmanyam [53,54], presented a simple methodology to design and implementation of Direct Torque Control (DTC) of three phase squirrel cage induction motor using Matlab/Simulink and FPGA software. Two simple new techniques i.e. constant switching frequency and stator flux estimation are proposed to maintain this simple control structure of DTC while at the same time improving the performance of the DTC drives. A simple torque control is introduced to replace the three level hysteresis comparators to maintain a constant switching frequency. By using simple compensator based on steady state operation, the magnitude and phase error associated with stator flux estimation based on voltage model is compensated.

In 2008, Pandya S.N, et.al [55], presented the basic issues associated with the direct torque control of induction motor drives. This technique mainly focuses to reduce the electromagnetic torque ripple using the concept of multi-rate sampling.

In 2008, Soliman, et.al, [56], proposed a method for improving the torque in PMSM drives using Simulink packages.

In 2010, Sanada, M., et.al, [57-63] proposed a control method suitable for permanent magnet synchronous motor drives for limited armature voltage and current based on direct torque. This method utilizes a mathematical approach in rotating reference frame synchronized to the stator flux linkage.

In 2010, Adballa, et.al, [64-66] proposed a control scheme for three phase induction motor using fuzzy logic controller to improve the performance and reducing the ripple in output torque. The dynamic performance of the model is enhanced and hence improves the operation.

In 2012, Amit Kumar, et.al, [67] proposed a simplified algorithm to reduce the computation time taken for switching of a three level inverter for synchronous and asynchronous machine.

In 2013, Han Di, et.al, [68] proposed the drive system to predict the stator resistance and flux linkage which was easily adaptable to rapid torque response.

In 2013, Begam, et.al, [69] proposed the implementation of a high performance direct torque control of induction motor based on three level inverter drive by calculating the three level neutral point clamped inverter for high power application.

In 2013, Ramesh T., et.al, [70] proposed Mamdani type controller using triangular membership function based type-two fuzzy logic speed controller to improve the dynamic response of induction motor drive.

1.3 OBJECTIVE

The main objective of this project work is to develop a controller with superior dynamic response, fast torque response, having low inverter switching frequency, low harmonic losses and high efficiency. The DTC controller is the best alternative to fulfill all the above characteristics to be good a controller. So the objective here is to study the most advanced control method i.e. direct torque control using space vector pulse width modulation and investigate its performance characteristics. In this thesis induction motor drive control is performed using conventional approach. To remove the ripple in torque and excess stator current two different approach are presented namely DTC using space vector pulse width modulation and fuzzy logic control.

1.4 ORGANIZATION OF THE THESIS

Chapter 2 discusses the mathematical model of three phase induction motor, the concept of reference frames and the induction motor dynamic equations in various reference frames, the equations in stationary reference frame are to be applied on further chapters.

Chapter 3 introduces the conventional DTC scheme, discusses DTC principle and various methods of estimating stator flux in detail, a brief idea about the hysteresis controllers is given.

Chapter 4 discusses the basic concept of space vector modulation, pulse pattern generation and applications of space vector pulse width modulation technique to indirect vector control strategy.

Chapter 5 discusses the DTC scheme using fuzzy logic control.

Chapter 6 Conventional DTC scheme is simulated by MATLAB/SIMULINK platform. The model is run for typical conditions of reference speed and applied torque values. Characteristics of motor for classical direct vector control and DTC based on SVPWM are shown.

Chapter 7 gives summery of the whole work, the conclusion and directions for future work.

CHAPTER 2

INDUCTION MOTOR DRIVES AND CONTROL TECHNIQUE

2.1 INTRODUCTION

Electric drives deals with ac waveforms, hence power converters required in adjustable speed drives should produce ac from dc supply. The sinusoidal parameters like magnitude, frequency, and phase should be controllable. Voltage source inverters are mostly used inverters in adjustable speed drives in industrial applications. The output of the inverter is fed to the motor which is further connected to the load.

2.2 DYNAMIC MODEL OF INDUCTION MOTOR

The stator of three phase induction motor consists of three windings separated in space by 120 degree. When the winding is connected to three phase supply a rotating magnetic field is set up in the stator and rotor by induction principle. For the study purpose squirrel cage motor is considered because of several advantages over slip ring induction motor [4,10]. Squirrel cage induction motors are more rugged in construction and free from slip rings, so less maintenance is required. The dynamic model of machine is considered for adjustable drive system. The dynamic behavior of induction machine is complex in nature as the rotor winding keeps on moving with respect to stator winding. Coupling coefficient between rotor and stator also changes with the rotor position.

Following assumptions are taken into account for dynamic model of the machine:

- Saturation effect of the magnetic core is neglected

- Saliency is neglected, so that the inductance offered by the winding is free from rotor position
- Stator windings are arranged in space to produce sinusoidal mmf
- Fringing effect is neglected
- Radial magnetic flux density across the air gap
- Eddy current and hysteresis effect is neglected

The per phase equivalent circuit of IM is only valid in steady state condition. In adjustable speed drives, due to the presence of feedback loop transient behavior of machine is required [15]. For better understanding dynamic d-q model is required. For dynamic modeling of motor two axis theory is used. According to this theory time varying parameters can be represented in two mutually perpendicular axes, direct axis (d-axis) and quadrature axis (q-axis). A three phase machine can be represented by equivalent two phase machine.

Where d^s-q^s represent stator direct and quadrature axis and d^r-q^r represents rotor direct and quadrature axis.

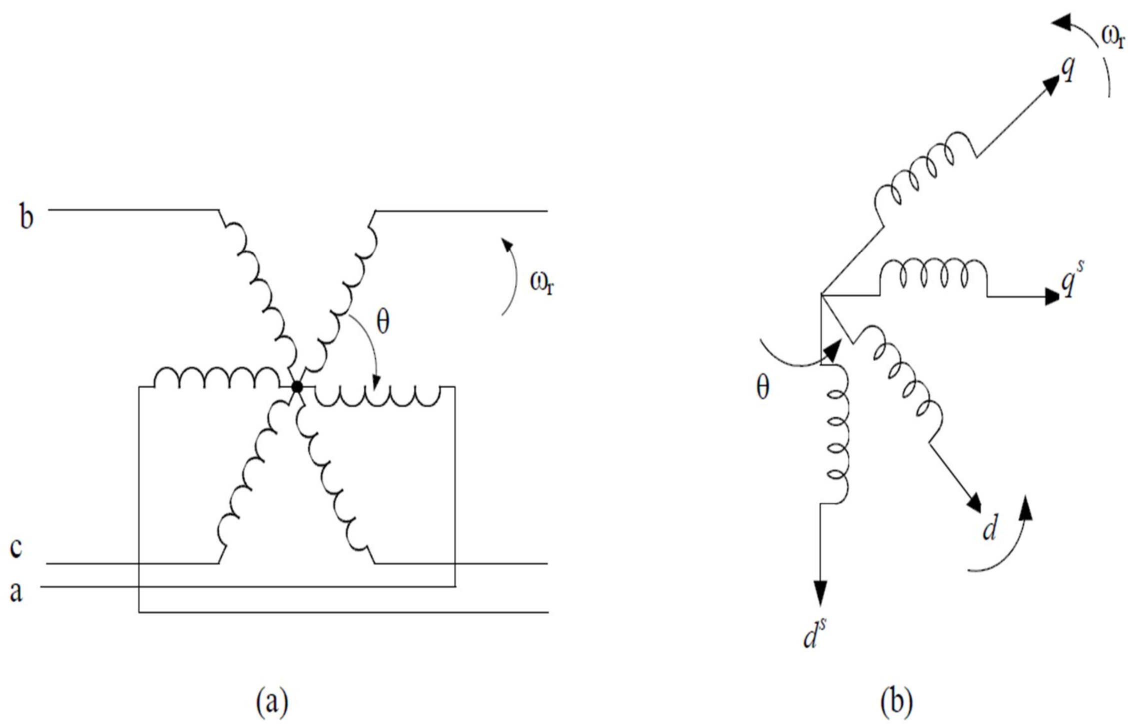

Fig 2.1 (a) coupling effect in stator and rotor winding of motor (b) equivalent two phase machine

2.3 AXIS TRANSFORMATION

2.3.1 Three- Phase to 2- Phase transformation:

A symmetrical three phase machine is considered with stationary axis as-bs-cs at 120 degree apart in space.

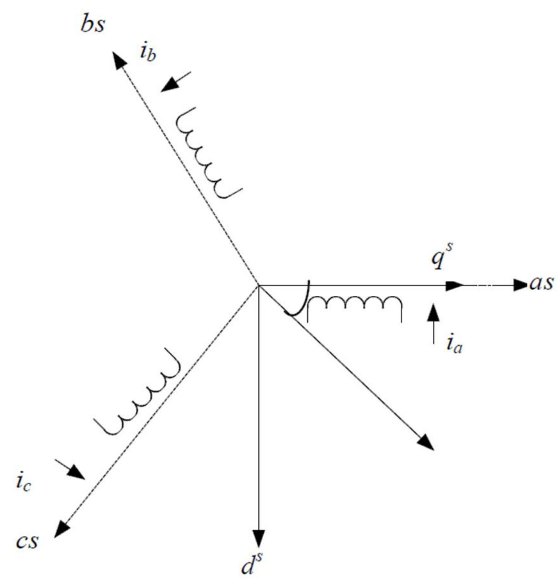

Figure 2.2 *as-bs-cs* to *ds-qs* axis transformation (θ = 0)

Voltages v_{as}, v_{bs}, v_{cs} are the voltages of *as, bs, cs* phases respectively. Let us consider the stationary *ds-qs* axes are oriented at θ angle as shown and the voltages along *ds-qs* axes to be v_{ds}^s and v_{qs}^s respectively, the stationary two phase voltages can be converted to three phase voltages according to the following equations:

$$v_{as} = v_{qs}^s \cos\theta + v_{ds}^s \sin\theta \tag{2.1}$$

$$v_{bs} = v_{qs}^s \cos(\theta - 120) + v_{ds}^s \sin(\theta - 120) \tag{2.2}$$

$$v_{cs} = v_{qs}^s \cos(\theta + 120) + v_{ds}^s \sin(\theta + 120) \tag{2.3}$$

Similarly two axis (*d-q*) voltages can be transformed to three phase symmetrical voltages as per the following equations:

$$v_{as} = v_{qs}^{s} \tag{2.4}$$

$$v_{bs} = -\frac{1}{2}v_{qs}^{s} - \frac{\sqrt{3}}{2}v_{ds}^{s} \tag{2.5}$$

$$v_{cs} = -\frac{1}{2}v_{qs}^{s} + \frac{\sqrt{3}}{2}v_{ds}^{s} \tag{2.6}$$

$$v_{qs}^{s} = v_{as} \tag{2.7}$$

$$v_{ds}^{s} = -\frac{1}{\sqrt{3}}(v_{bs} - v_{cs}) \tag{2.8}$$

2.3.2 Two phase stationary to two phase synchronously rotating frame transformation:

The stationary *ds-qs* axes are transformed to synchronously rotating *de-qe* reference frame which is rotating at speed ωe with respect to *ds-qs* axes with the help of fig.2.3.The angle between *ds* and *de* axes is $\theta_e = \omega_e t$. The voltagesv_{ds}^{s} and v_{qs}^{s} can be converted to voltages on *de-qe* axis according to the following relations:

$$v_{ds} = v_{qs}^{s}\cos\theta_s - v_{ds}^{s}\sin\theta_s \tag{2.9}$$

$$v_{qs} = v_{qs}^{s}\sin\theta_s + v_{ds}^{s}\cos\theta_s \tag{2.10}$$

The transformation of rotating frame parameters to stationary frame is according to the following relations:

$$v_{qs}^{s} = v_{qs}\cos\theta_e + v_{ds}\sin\theta_e \tag{2.11}$$

$$v_{ds}^{s} = -v_{qs}\sin\theta_e + v_{ds}\cos\theta_e \tag{2.12}$$

Assuming that the three phase voltages are balanced and sinusoidal given by following

$$v_{as} = v_m\cos(\omega_s t + \Phi) \tag{2.13}$$

$$v_{bs} = v_m\cos(\omega_s t + \Phi - {}^{2\pi}/_{3}) \tag{2.14}$$

$$v_{cs} = v_m \cos\left(\omega_s t + \Phi - \frac{2\pi}{3}\right) \quad (2.15)$$

Substituting equations (2.13) – (2.15) in equations (2.8) and (2.9) we get

$$v_{qs}^{s} = v_m \cos(\omega_e t + \Phi) \quad (2.16)$$

$$v_{ds}^{s} = v_m \sin(\omega_e t + \Phi) \quad (2.17)$$

Substituting (2.16) – (2.17) in (2.9) – (2.10)

$$v_{qs} = v_m \cos\Phi \quad (2.18)$$

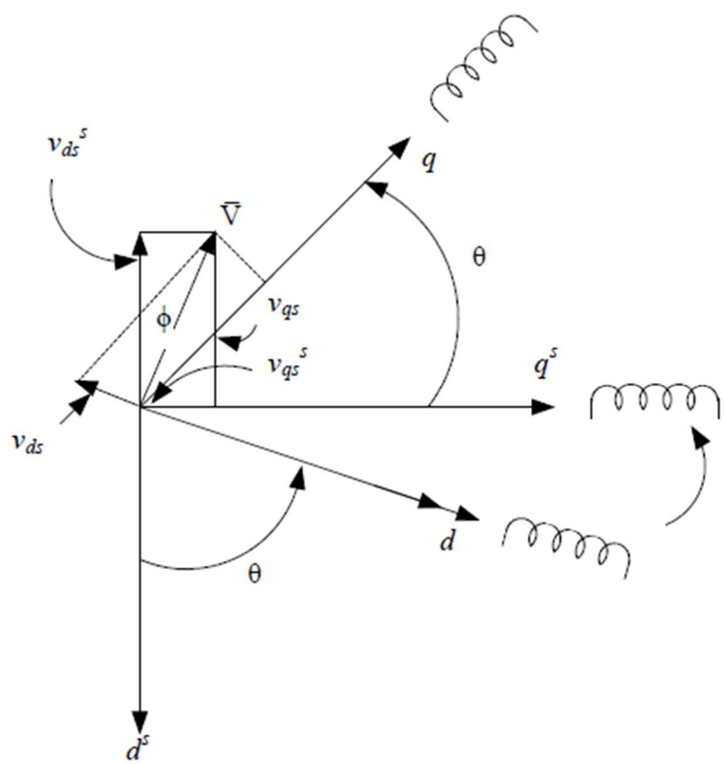

Figure 2.3 Stationary *d-q* frame to synchronously rotating frame transformation

$$v_{ds} = -v_m \sin\Phi \quad (2.19)$$

From equations (2.18) and (2.19) it is clear that sinusoidal quantities in a stationary frame appear as dc quantities in a synchronously rotating reference frame.

2.4 MOTOR DYNAMIC MODEL IN STATIONARY FRAME

Machine model in stationary frame by Stanley equations substituting $\omega_e = 0$. The stator circuit equations are written as:

$$v_{qs}^{s} = R_s i_{qs}^{s} + (d/_{dt})\Psi_{qs}^{s}) \tag{2.20}$$

$$v_{ds}^{s} = R_s i_{ds}^{s} + (d/_{dt})\Psi_{ds}^{s} \tag{2.21}$$

$$0 = R_r i_{qr}^{s} + \left(d/_{dt}\right)\Psi_{qr}^{s} - \omega_r \Psi_{dr}^{s} \tag{2.22}$$

$$0 = R_r i_{dr}^{s} + \left(d/_{dt}\right)\Psi_{dr}^{s} + \omega_r \Psi_{qr}^{s} \tag{2.23}$$

Where,

$\Psi_{qs}^{s}, \Psi_{ds}^{s}$ = q-axis and d-axis stator flux linkages

Ψ_{qr}^{s} Ψ_{dr}^{s}= q-axis and d-axis rotor flux linkages

R_s ,R_r= stator and rotor resistances

ω_r = rotor speed and v_{dr}= v_{qr} = 0

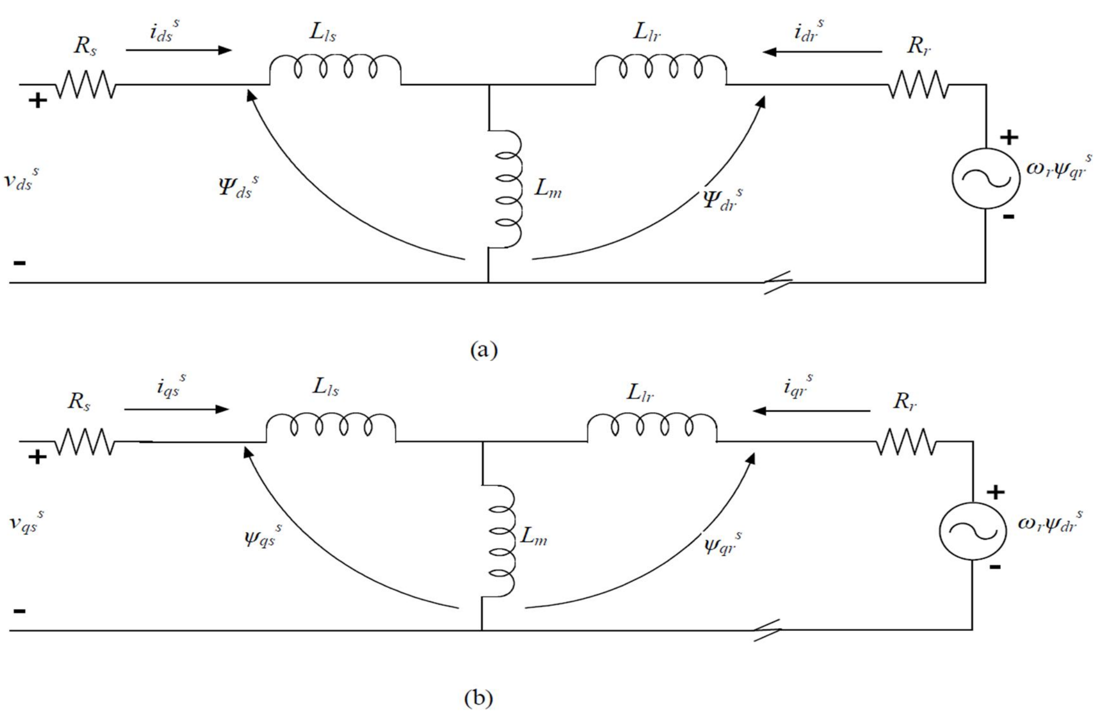

Figure 2.4(a),(b) d^s-q^s equivalent circuits

The electromagnetic torque is developed by the interaction of air gap flux and rotor mmf which can be expressed in general vector form as

$$T_e = \frac{3}{2}(\frac{P}{2})(\overline{\Psi_m})\ (\overline{I_r}) \tag{2.24}$$

The torque equations can be written in stationary frame with corresponding variables as

$$T_e = (\frac{3}{2})(\frac{P}{2})(\Psi_{dm}^{s} i_{dr}^{s} - \Psi_{qm}^{s} i_{dr}^{s}) \tag{2.25}$$

$$= (\frac{3}{2})(\frac{P}{2})(\Psi_{dm}^{s} i_{qs}^{s} - \Psi_{qm}^{s} i_{ds}^{s}) \tag{2.26}$$

$$= (\frac{3}{2})(\frac{P}{2})(\Psi_{ds}^{s} i_{qs}^{s} - \Psi_{qs}^{s} i_{ds}^{s}) \tag{2.27}$$

$$= (\frac{3}{2})(\frac{P}{2})L_m(i_{dr}^{s} i_{qs}^{s} - i_{qr}^{s} i_{ds}^{s}) \tag{2.28}$$

$$= (\frac{3}{2})(\frac{P}{2})(\Psi_{dr}^{s} i_{qr}^{s} - \Psi_{qr}^{s} i_{qr}^{s}) \tag{2.29}$$

2.5 INDUCTION MOTOR CONTROL

There are various methods available for control of Induction Motor. These are:

- Open loop volts/hertz control
- Closed loop speed control with sleep compensation
- Flux oriented or vector control
- Direct torque control
- Adaptive control
- Artificial intelligence based control

2.5.1 Open loop voltage/hertz control:

In AC machine stator flux is equal to the stator voltage to frequency ratio since the motor is fed with a variable AC source voltage and frequency, it is important to maintain the (V/F) ratio constant in the constant torque region if there is no magnetic saturation [29].

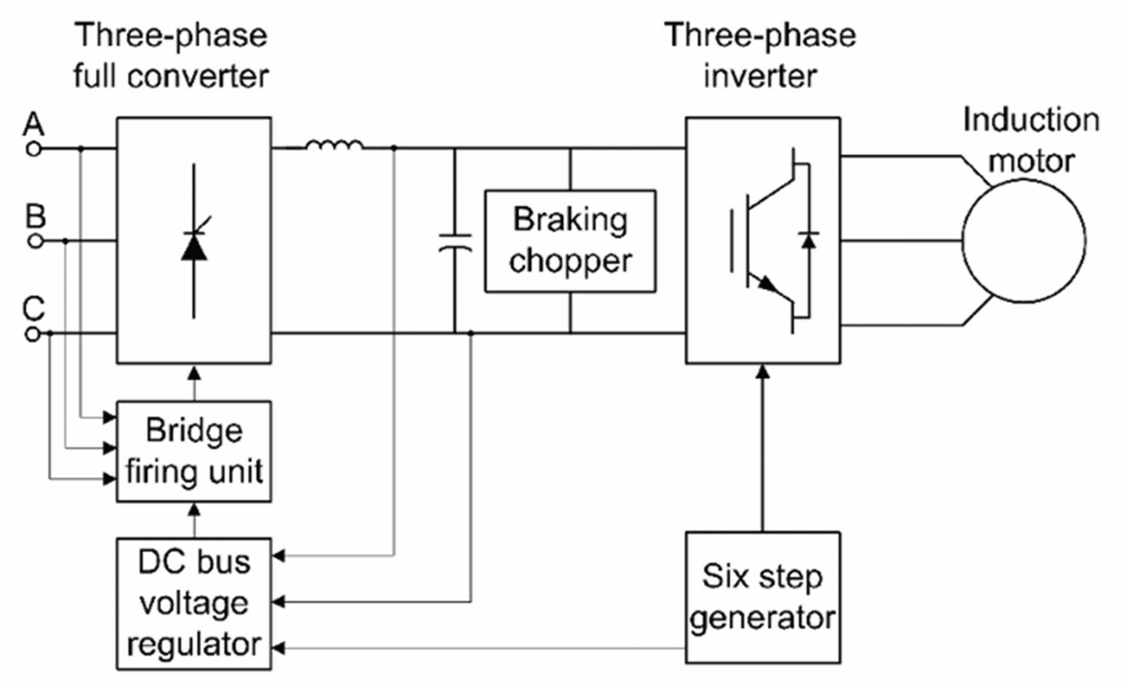

Figure 2.5 block diagram of open loop voltage/hertz control

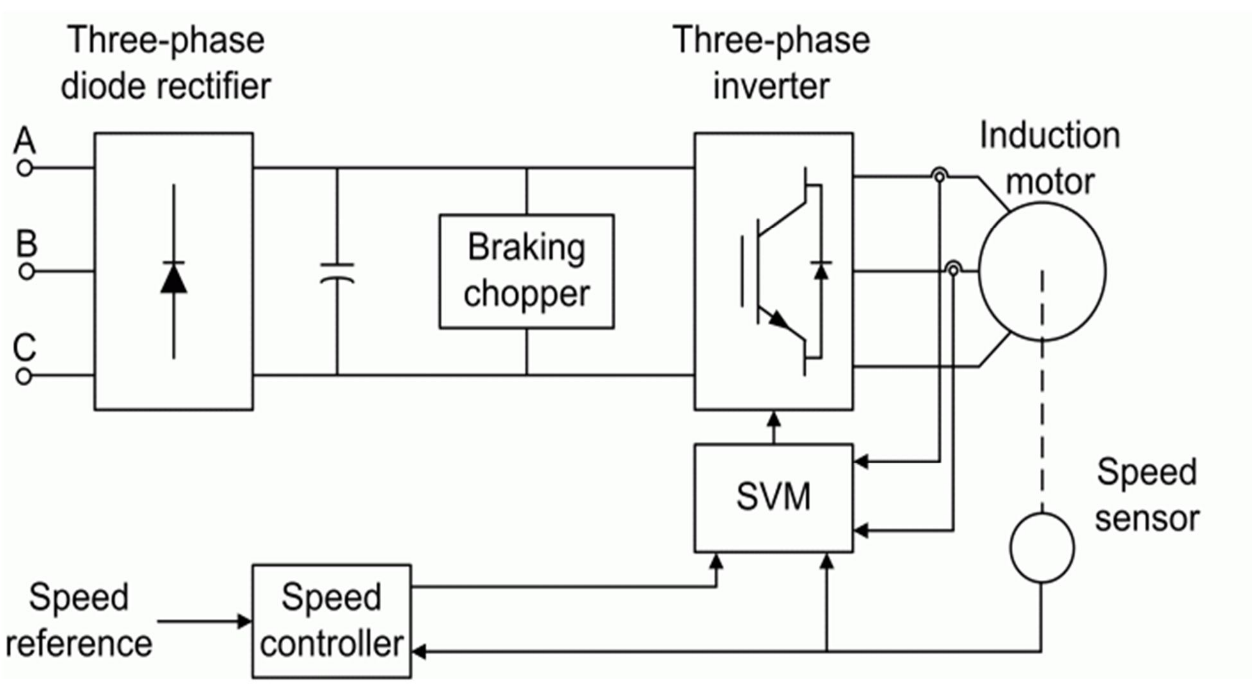

Figure 2.6 block diagram of closed loop control

2.5.2 Closed loop speed control with sleep compensation:

In this type of control, a slip speed command is combined with the measured rotor speed to produce the required inverter frequency. Slip command is obtained from a PI based regulator. The schematic of the closed loop speed control with sleep compensation, shown below consists of six main blocks, induction motor, three-phase inverter, three-phase diode rectifier, speed controller, braking chopper, and space vector modulator.

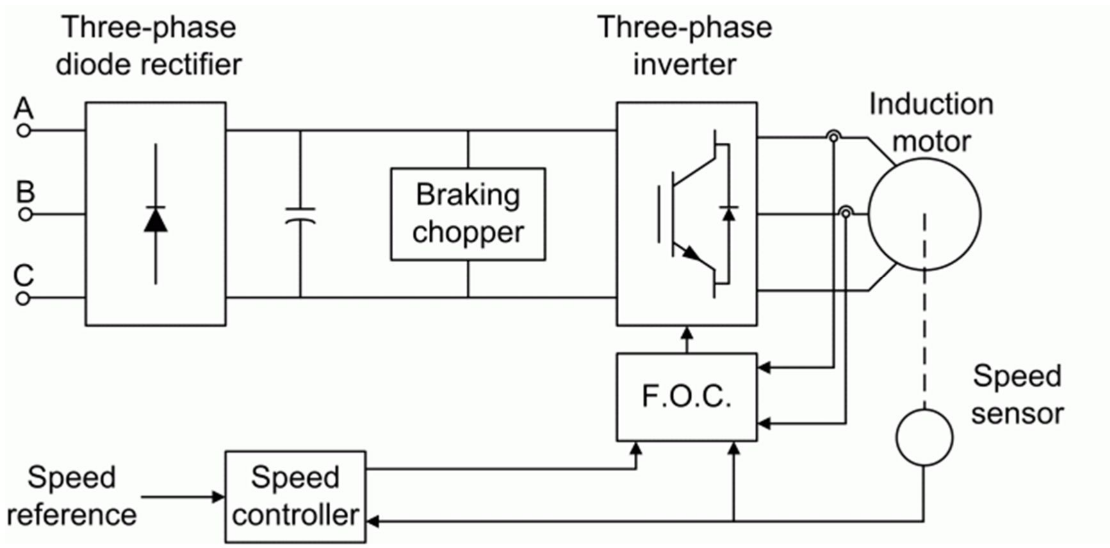

Figure 2.7 Block diagram of field oriented control scheme

2.5.3 Flux oriented control:

In AC machine stator and the rotor fields are not orthogonal but vary with the operating conditions. DC machine-like performance can be obtained by holding a fixed and orthogonal orientation between the field and armature fields in an AC machine by maintaining the stator current with respect to the rotor flux so as to attain independently controlled flux and torque. This type of control is known as flux oriented control or vector control [20-23].

2.5.4 Direct Torque Control:

Field oriented control is a good control technique but the major drawback is that it depends on accurate knowledge of the motor parameters [17-19]. A more attractive control method consists first in removing the machine stator flux and electric torque in the stationary reference frame from terminal measurements. The following relations are used:

$$\Psi_{ds}^{s} = (v_{ds}^{s} - i_{ds}^{s} R_s)dt \tag{2.30}$$

$$\Psi_{qs}^{s} = \left(v_{qs}^{s} - i_{qs}^{s} R_s\right)dt \tag{2.31}$$

$$\Psi_{s} = (\Psi_{ds}^{s^2} + \Psi_{qs}^{s^2}) \tag{2.32}$$

2.6 Conclusion

In this chapter detail dynamic model of IM is discussed in stationary reference frame. In order to understand and design vector controlled drives the dynamic model of the machine to be controlled must be known which could be a good approximation of the real plant. To formulate the model two axis theory and space phasor notations have been used. It has been proved that space phasor notation is compact and easier to work with.

CHAPTER 3

DIRECT TORQUE CONTROL OF INDUCTION MOTOR

3.1 INTRODUCTION

In recent years “induction motor control techniques” have been the field of interest of many researchers to find out various solutions for induction motor control having the features of more accurate and faster torque response, and reduction of the complexity of field oriented control. The Direct torque control (DTC) technique has been recognized as the simple and viable solution to achieve these requirements. DTC is one of the most excellent and efficient control strategies of induction motor [10-12]. This technique is based on decoupled control of torque and stator flux and today it is one of the most actively researched control techniques where the aim is to control effectively the torque and flux.

3.2 CONVENTIONAL DTC SCHEME

The conventional DTC method is a closed loop control method, the blocks of the control structure are: the power supply unit, a three phase voltage source inverter unit, the induction motor, a speed controller block to generate the torque command and the DTC controller. The DTC controller further consists of torque and flux estimation block, two hysteresis controllers and voltage vector sector selection block, the output of the DTC controller is the gating pulses for the inverter. The DTC scheme do not need coordinate transformation because all the control procedures are carried out in stationary frame of reference [27,28]. Hence this method does not suffer from parameter variations as compared to other control techniques. Also there is no feedback current control loop due to which the control actions do not suffer from the delays inherent in the current controllers, no pulse width modulator, no PI controllers, and no rotor speed or position sensor. So it is a sensor less control technique which operates the motor without requiring a shaft mounted mechanical sensor. Here on-line torque and flux estimators are used for closing the loop. Here the torque and stator flux are controlled directly by using hysteresis comparators. Fig.3.1 shows the basic block diagram of conventional DTC scheme.

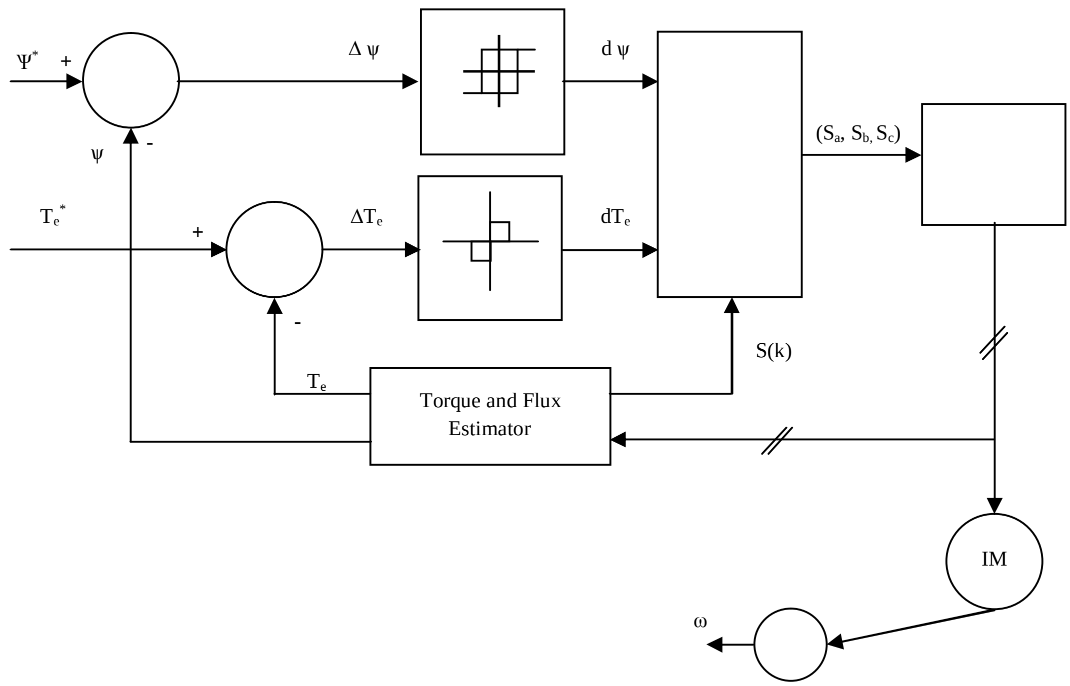

Figure 3.1: Block Diagram of Conventional DTC scheme for Induction Motor Drives

3.3 PRINCIPLE OF DTC SCHEME

The basic principle of DTC is to directly select stator voltage vectors according to the torque and flux errors which are the differences between the references of torque and stator flux linkage and their actual values. The governing equation for torque for this scheme is due to the interaction of stator and rotor fields. Torque and stator flux linkages are computed from measured stator voltages and current. An optimal voltage vector for the switching of VSI is selected among the six nonzero voltage vectors and two zero voltage vectors by the hysteresis control of stator flux and torque [30-31].

A three-phase VSI has eight possible combinations of six switching devices which is shown in Figure 3.2. The six switches have a well-defined state: ON or OFF in each configuration. So all the possible configurations can be identified with three bits (S_a, S_b, S_c), one for each inverter leg. The bit is set to 1 if the top switch is closed and to 0 when the bottom switch is closed. In order to prevent short circuit of the supply, the state of the upper switch is always opposite to that of the lower one.

The stator voltage space vector is

$$\bar{V_S} = \frac{2}{3}E[S_a + e^{j\frac{2\pi}{3}}S_b + e^{j\frac{2\pi}{3}}S_c] \tag{3.1}$$

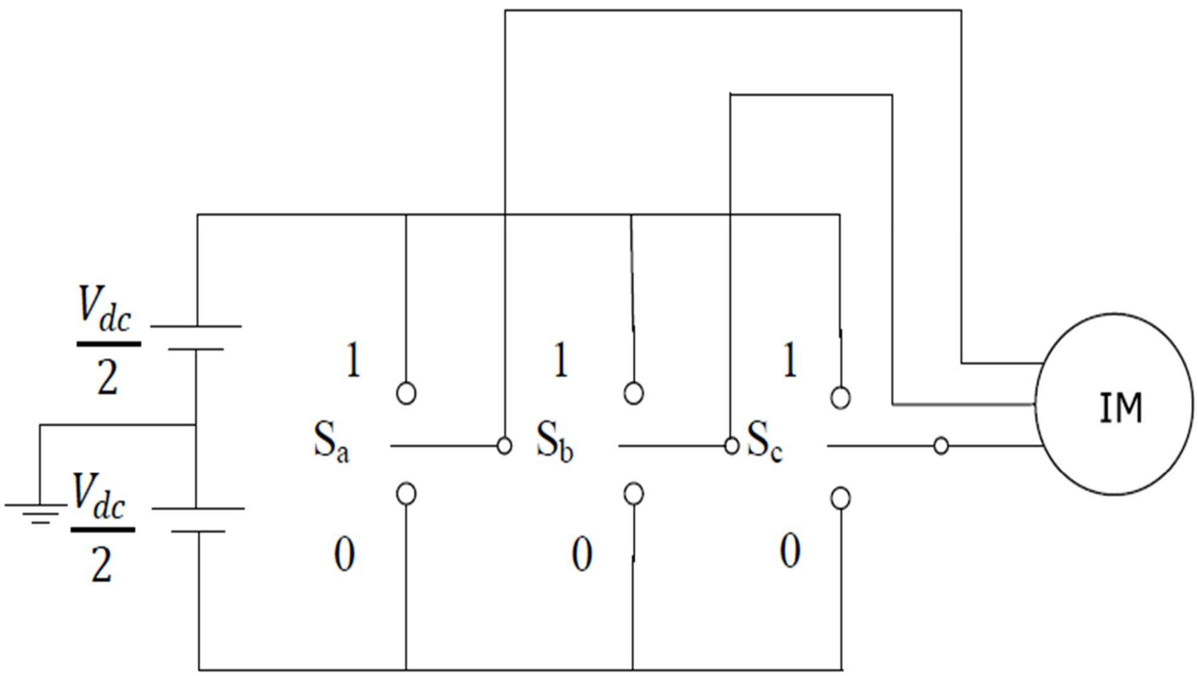

Figure 3.2 Three Phase VSI

3.4 DIRECT FLUX CONTROL

In stationary reference frame the stator flux equation can be written as:

$$\Psi_s = \int(\bar{v_s} - \bar{\iota_s}R_s)dt \tag{3.2}$$

If the stator resistance drop is neglected for simplicity, the stator flux varies along the direction of applied voltage vector and the equation will be reduced to

$$\Delta\overline{\Psi_s} = \bar{v_s}\Delta t \tag{3.3}$$

Which means, by applying stator voltage vector *Vs* for a time increment *Δt, Ψs* can be changed incrementally. The command value of the stator flux vector *Ψs** follows a circular trajectory, the

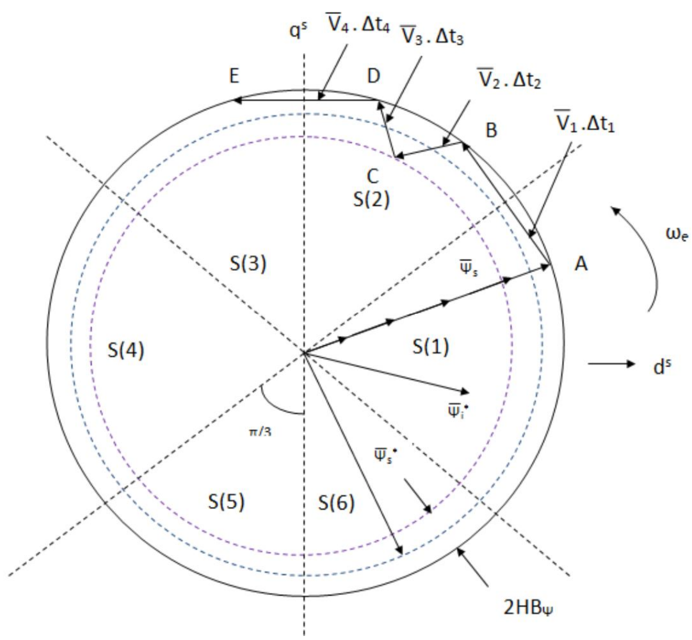

Figure 3.3 Circular trajectory of stator flux

plane of stator flux is divided into six sectors as shown in fig.3.3. Each sector has a different set of voltage vector to increase or decrease the stator flux. The command flux vector rotates in anticlockwise direction in a circular path and the actual stator flux vector *Ψs* tracks the command flux in a zigzag path but constrained to the hysteresis band which is shown in Figure 3.3.

In general the active forward voltage vectors (*Vs,k+1* and *Vs,k+2*) are applied to increase or decrease the stator flux respectively when the stator flux lies in sector *k*. The radial voltage vectors (*Vs,k* and *Vs,k+3*) which quickly affect the flux are generally avoided. The active reverse voltage vectors (*Vs,k-1* and *Vs,k-2*) are used to increase or decrease the stator flux in reverse direction [6-7].

The stator flux vector change due to stator voltage vector is quick whereas change rotor flux is sluggish because of its large time constant *Tr*. That is why *Ψs* movement is jerky and *Ψr* moves uniformly at frequency ωe as it is more filtered. However the average speed of both remains the same in steady state condition.

3.5 DIRECT TORQUE CONTROL

The electromagnetic torque produced due to interaction of stator and rotor flux is given by the following equation:

$$T_e = \frac{3}{2}\left(\frac{P}{2}\right)\frac{L_m}{L_s' L_r}\overline{\Psi_s^* \Psi_r} = \frac{3}{2}\left(\frac{P}{2}\right)\frac{L_m}{L_s' L_r}\Psi_s \Psi_r \sin\gamma \quad (3.4)$$

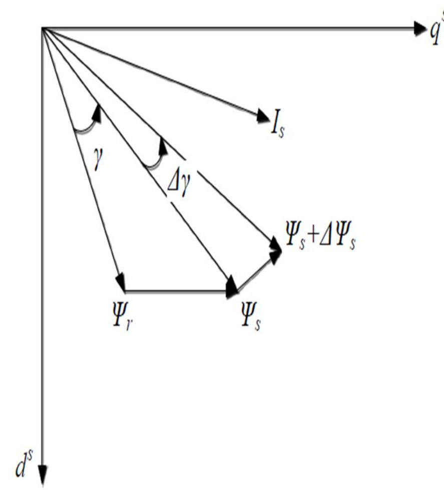

Figure.3.4 Stator flux, rotor flux and stator current vectors in d^s-q^s reference plane

From the above it is clear that torque varies directly as angle between stator flux and rotor flux i.e. γ. So in order to obtain high dynamic performance it is required to vary γ quickly. Assuming the rotor is rotating in anticlockwise direction continuously and stator flux lies in sector *k*, the active forward voltage vectors (*Vs, k+1* and *Vs, k+2*) are applied to increase γ so as the torque *Te*. The radial voltage vectors (*Vs, k* and *Vs, k+3*) are used to decrease γ and *Te*. By applying the reverse active voltage vectors (*Vs,k-1* and *Vs,k-2*) torque can be decreased rapidly. The two zero voltage vectors (*Vs,0* and *Vs,7*) are applied to maintain the flux constant ideally and to decrease the torque slightly.

3.6 SWITCHING SELECTION

A high performance torque control can be established due to the decoupled control of stator flux and torque in DTC. Fig.3.5 shows an example of stator flux located in sector-1 (S(1)) with the corresponding optimum switching voltage vectors for anti-clockwise and clockwise rotation of the shaft.

Optimum switching vector selection table given by table 3.1 shows the optimum selection of the switching vectors in all sectors of the stator flux plane. This table is based on the value of stator flux error status, torque error status and orientation of stator flux for counterclockwise rotation of the shaft.

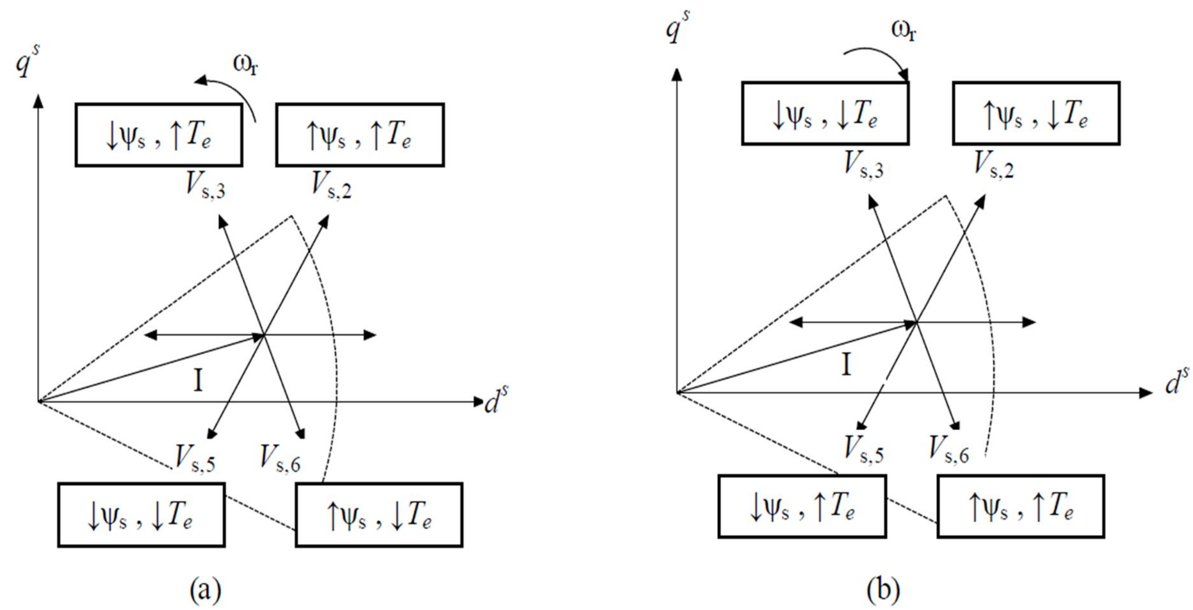

Figure3.5 Optimum switching voltage vector in sector-1 for (a) anti-clockwise and (b) clockwise Rotation

$d\psi$	dTe	S(1)	S(2)	S(3)	S(4)	S(5)	S(6)
1	1	V_2	V_3	V_4	V_5	V_6	V_1
	0	V_7	V_0	V_7	V_0	V_7	V_0
	-1	V_6	V_1	V_2	V_3	V_4	V_5
0	1	V_3	V_4	V_5	V_6	V_1	V_2
	0	V_0	V_7	V_0	V_7	V_0	V_7
	-1	V_5	V_6	V_1	V_2	V_3	V_4

Table 3.1: Applied selected voltage vectors

3.7 STATOR FLUX ESTIMATION

For exact calculation of stator flux and torque errors, an accurate estimator of stator flux is necessary. There are commonly used methods of estimation of flux namely stator voltage model and current model.

3.8 STATOR VOLTAGE MODEL

This is the simplest method of stator flux estimation, where the machine terminal voltages and currents are sensed and from the stationary frame equivalent circuit the fluxes are computed. The estimated stator flux is given by the following equation:

$$\Psi_{ds}^{s} = \int (v_{ds}^{s} - i_{ds}^{s} R_s) dt \quad (3.5)$$

$$\Psi_{qs}^{s} = \int \left(v_{qs}^{s} - i_{qs}^{s} R_s\right) dt \quad (3.6)$$

$$\Psi_s = \sqrt{(\Psi_{qs}^{s^2} + \Psi_{qs}^{s^2})} \quad (3.7)$$

This method provides accurate flux estimation at high speed but in industrial applications requiring vector drives at zero start-up this method cannot be used because at low speed stator resistance drop becomes significant causing inaccurate estimation. Also at low frequency, voltage signals are very low and dc offset tends to build up at the integration output, as a result ideal integration becomes difficult.

3.9 CURRENT CONTROL

In stationary reference frame, current model is globally stable and the drives operation can be extended down to zero speed. But this model is much complex as compared to voltage model as here the knowledge of rotor speed and stator current is required to estimate rotor flux linkage and stator flux can be estimated based on the estimation of rotor flux linkage [34]. From the dynamic equations of IM in stationary reference frame, stator and rotor flux can be derived which are given below:

$$\frac{d}{dt}\overline{\Psi_r} = \left(\frac{L_m i_s - \Psi_r}{T_r}\right) - \omega_r \Psi_r \quad (3.8)$$

$$\overline{\Psi_s} = \frac{L_m}{L_r}\overline{\Psi_r} + \sigma L_s i_s \quad (3.9)$$

Here the equations involve closed loop integration, so there is no integration drift problem in current model at low speed region. However estimation accuracy is affected due to motor parameter variation, particularly rotor resistance variation becomes dominant by skin effect and temperature.

It is ideal to have a hybrid model based on the unique features gained by both models respectively where the voltage model would be effective at higher speed range and current model at lower speed range.

3.10 HYSTERESIS CONTROLLER

DTC of induction motor drives requires two hysteresis controllers. The drive performance is influenced by the width of the hysteresis bands in terms of flux and torque ripples, current harmonics and switching frequency of power electronics devices. Current distortion is reduced by small flux hysteresis band and torque ripple is reduced by small torque hysteresis bands. In each sampling time, the switching state of the inverter is updated [37]. The inverter state remains constant, until the output states of the hysteresis controller change within a sampling interval. If the hysteresis band is fixed, the switching frequency totally depends on the rate of change of torque and flux.

3.11 TORQUE HYSTERESIS CONTROLLER

The Torque hysteresis controller is a three level controller. It means the torque control loop has three levels of digital outputs. The torque error ΔTe is given to the torque hysteresis controller and the output is torque error status (dTe) which can have three values -1, 0 or 1. The width of the hysteresis band is 2ΔTe. Torque error status is given to the switching table for optimum voltage vector selection for the inverter.

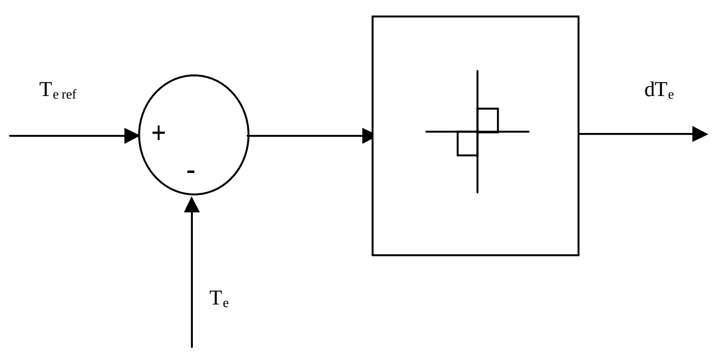

Figure 3.6: Torque Hysteresis Control

Torque error $\Delta Te = Teref - Te$

$|dTe| = 1$ if $|Te| < |Teref| - |\Delta Te|$: Torque to be increased

$|dTe| = -1$ if $|Te| > |Teref| + |\Delta Te|$: Torque to be decreased

$|dTe| = 0$ if $|Teref| - |\Delta Te| \leq |Te| \leq |Teref| + |\Delta Te|$: Torque to remain unchanged

3.12 FLUX HYSTERESIS CONTROLLER

The flux hysteresis controller is a two level controller. So the flux control loop has two digital outputs. The stator flux error $\Delta\psi s$ is given to the flux hysteresis controller and the output is flux error status ($d\psi s$) which can have two values 0 and 1. The width of the hysteresis band is $2\Delta\ \psi s$. Flux error status is given to the switching table for optimum voltage vector selection for the inverter.

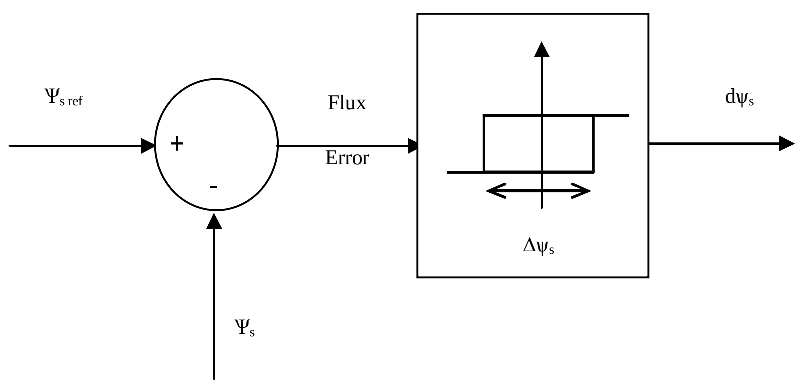

Figure 3.7: Flux Hysteresis Control

Stator flux error $\Delta\psi_s = \psi_{sref} - \psi_s$

The flux is controlled according to the following equations

$|d\psi_s| = 1$ if $|\psi_s| \leq |\psi_{sref}| - |\Delta\psi_s|$:flux to be incrased

$|d\psi_s| = 0$ if $|\psi_s| \geq |\psi_{sref}| + |\Delta\psi_s|$: flux to be decreased

3.13 Conclusion

This chapter included the detail description of the conventional DTC scheme and principle of indirect FOC scheme. This DTC scheme has many advantages over field oriented control which has been discussed in this chapter but also has some drawbacks like generation of flux and torque ripple and variable switching frequency. The flux and torque ripple is due to the hysteresis controller which can however be reduced significantly by reducing the sampling period. The variable switching frequency is due to sector change of the stator flux vector.

CHAPTER 4

SPACE VECTOR PULSE WIDTH MODULATION FED INDUCTION MOTOR DRIVES

4.1 INTRODUCTION

The use of PWM drive is advantageous in many ways, for example it obtains its dc input through uncontrolled rectification of commercial AC mains and has good power factor, good efficiency, relatively free from regulation problems, it has the ability to operate the motor with nearly sinusoidal current waveform. The conventional PWM techniques are suitable for open loop control, for the implementation of a closed loop controlled AC drive Space vector PWM (SVPWM) technique is applied [8]. In this technique, the switching patterns for the bridge inverter are generated from the knowledge of stator voltage space phasor. A reference voltage vector is generated to generate a field synchronous with the rotating voltage vector by utilizing the different switching states of a three phase bridge inverter.

4.2 THEORY OF SVPWM

When three phase supply is given to the stator of the induction machine, a three phase rotating magnetic field is produced. Due to this field flux, a three phase rotating voltage vector is generated which lags the flux by 90º. This field can also be realized by a logical combination of the inverter switching which is the basic concept of SVPWM [16].

4.3 VOLTAGE SPACE PHASOR

The three phase bridge inverter has eight possible switching states: six active and two zero states. The six switches have a well-defined state ON or OFF in each configurations. At a particular instant, only one switch in each of the three legs is ON. Corresponding to each state of the inverter, there is one voltage space vector. For example for state zero it is V_0, for state 1 it is V_1 and so on. These switching state vectors have equal magnitude but 60º apart from each other. These vectors can be written in generalized form as follows:

$$v_k = v_{dc} e^{j(\frac{k-1}{3})\pi} \quad ; k=1,2,\ldots\ldots6 \tag{4.1}$$

$$= 0 \quad ; k= 0,7$$

Where

k = inverter state number.

v_{dc} = dc link voltage of the inverter

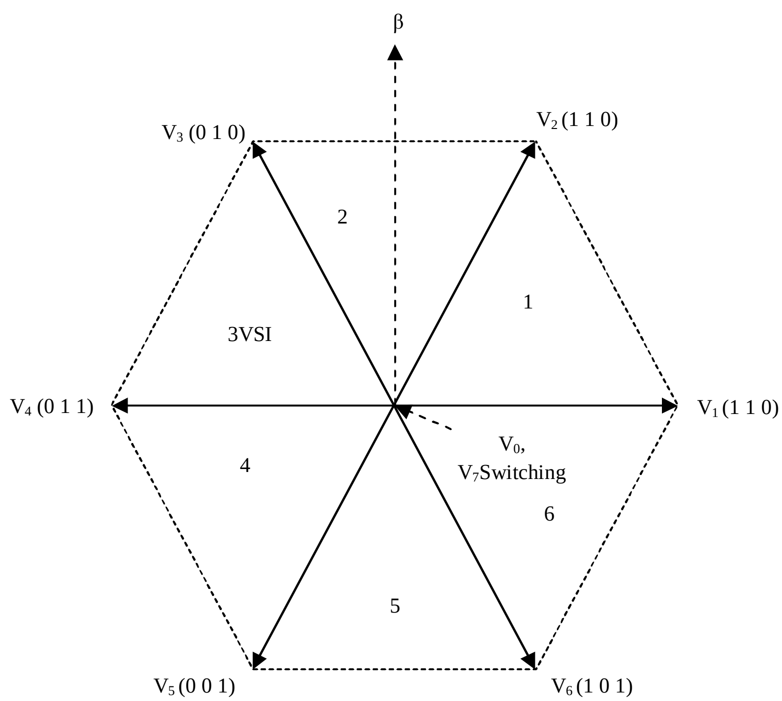

Figure 4.1: Inverter Switching State Vectors

The space bounded by two inverter space vectors is called a sector. So the plane is divided into six sectors each spanning 60º. In a balanced three phase system the voltage vectors are 120º apart in space and are represented by rotating vectors, whose projections on the fixed three phase axes are, sin0usoidal waves. So they can be represented as three sinusoidal references by a voltage reference space vector v_{ref}^{*} or v_s^*. The reference vector is assumed to be rotating in counter-clockwise direction with respect to d_s-axis (α-axis) as shown in fig.4.2 through six sectors.

The reference space vector can be synthesized by a combination of eight state vectors and is constant in magnitude at a switching instant t_s in case the switching frequency much higher than the output frequency. In a time average sense the reference vector at that instant can be approximated by two active voltage states of the inverter. For only certain amount of time these states are valid [21-24].

$$v_s^* = v_k t_k + v_{k+1} t_{k+1} \;\; ; \;\; k=0,1,2,\ldots\ldots\ldots,7 \tag{4.2}$$

In SVPWM, it is assumed that the space phasor of stator voltage v_s^*, is moving in α-β plane with constant angular velocity describing approximately a circular path. The basis of SVPWM scheme is to sample the v_s^* at sufficiently high rate, in between the sampling instants the vector is assumed to be constant in magnitude as shown in fig.4.2.

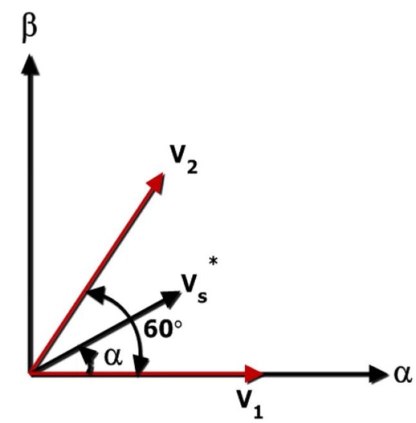

Figure 4.2 Reference vector in sector 1

In sector 1, the space voltage vector v_1 is along α-axis, v_2 makes an angle 60° to v_1 and at a particular instant v_s^* is making an angle γ w.r.to V_1. To generate the reference space vector in sector 1, the switching state vector v_1 is applied for an interval t_1, v_2 for t_2 and the two zero vectors v_0, v_7 for intervals t_0, t_7 respectively. So the total sampling interval t_s can be written as:

$$t_s = t_1 + t_2 + t_0 + t_7 \tag{4.3}$$

Resolving v_s^*, v_1 and v_2 along the α-β axis, and by equating voltage-time integrals we get:

$$|v_s^*| t_s \cos\gamma = |v_1| t_1 + |v_2| t_2 \cos\frac{\pi}{3} \quad (4.4)$$

$$|v_s^*| t_s \sin\gamma = |v_2| t_2 \sin\frac{\pi}{3} \quad (4.5)$$

Dividing both sides of equation (4.4) and (4.5) by V_{dc} and substituting (v_s^*/v_{dc})= a, we get:

$a t_s \sin\gamma = (\sqrt{3}/2)\, t_2$

Or $$t_2 = \frac{2}{\sqrt{3}} a t_s \sin\gamma \quad (4.6)$$

And $$a t_s \cos\gamma = t_1 + \frac{t_2}{2} \quad (4.7)$$

Substituting the value of t_2 from equation (4.6) in (4.7) and multiplying both sides of resulting equation by $\sqrt{3}/2$, we obtain

$$t_1 = \frac{2 a t_s}{\sqrt{3}} \sin\left(\frac{\pi}{3} - \gamma\right) \quad (4.8)$$

$$t_0 = t_7 = t_s - (t_1 + t_2) \quad (4.9)$$

Where

a = modulation index = (v_s^{*}/v_{dc})

By the knowledge of t_0, t_1, t_2, t_7 the switching pattern can be determined if the vector is in sector 1. The four time intervals change simultaneously when v_s^* goes from one sector to another for a particular modulation index a. The full cycle is completed by six similar sectors with label 1, 2,….,6. As v_s^* move over to sector 2, the inverter remains in switching state vector v_2 for time interval t_1 and in v_3 for time t_2. For sector 3: v_3 for t_1 and v_4 for t and so on.

4.4 PULSE PATTERN GENERATION

The PWM pattern generation means the generation of gating pulses for the six switches of the inverter, for correct interval so that appropriate switching state vectors are active for the appropriate time intervals as the reference space vector moves over a full cycle. In order to obtain minimum switching frequency, it is desired that only one phase of the inverter changes state from $+v_{dc}/2$ to $-v_{dc}/2$. while changing the switching vectors. So the arrangement of the switching sequence should be such that the transition from one state to the next state is performed by switching only one inverter phase. This is done by switching the inverter legs in a sequence starting from one zero state and ending at another zero state. Fig.4.3 shows the optimum inverter phase to dc center tap voltages. It can be noted that the switching frequency of the inverter is half of the sampling frequency [51-52].

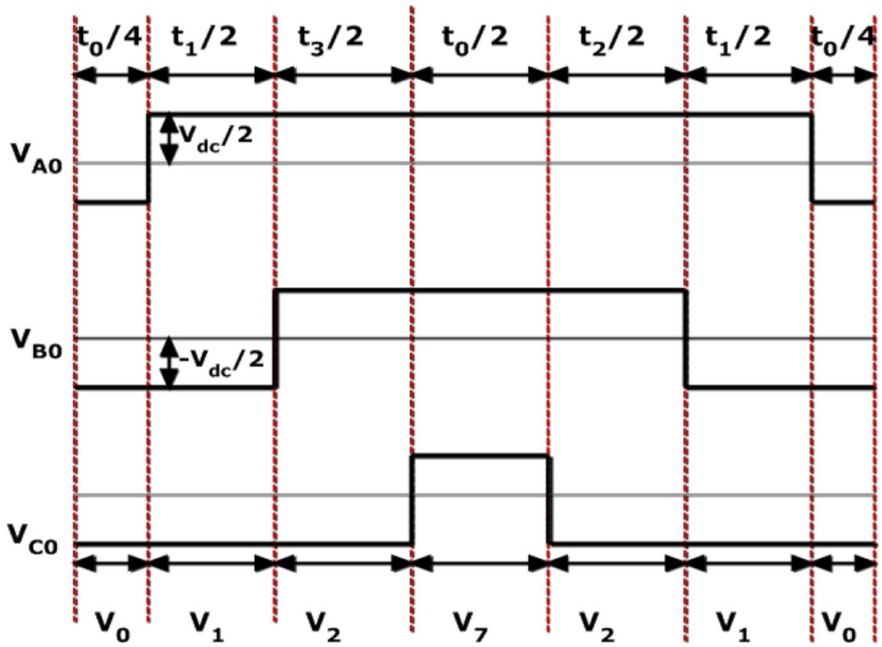

Figure 4.3 Leg voltages and space vector disposition in sector 1

The mean values of the phase to center tap voltages (V_{A0}, V_{B0}, V_{C0}) can be evaluated, averaging over one sampling period t_s as follows:

$$\overline{v_{A0}} = \frac{v_{dc}}{2t_s}(-\frac{t_0}{2} + t_1 + t_2 + \frac{t_0}{2}) \quad (4.10)$$

$$\overline{v_{B0}} = \frac{v_{dc}}{2t_s}(-\frac{t_0}{2} - t_1 + t_2 + \frac{t_0}{2}) \quad (4.11)$$

$$\overline{v_{C0}} = \frac{v_{dc}}{2t_s}(-\frac{t_0}{2} - t_1 - t_2 + \frac{t_0}{2}) \quad (4.12)$$

Substituting the values of t_1, t_2 in above three equations

$$\overline{v_{A0}} = \frac{av_{dc}}{\sqrt{3}}\sin(\gamma + \frac{\pi}{3}) \tag{4.13}$$

$$\overline{v_{B0}} = av_{dc}\sin(\gamma + \frac{\pi}{6}) \tag{4.14}$$

$$\overline{v_{c0}} = -\overline{v_{A0}} \tag{4.15}$$

4.5 CONCLUSION

The mean value of the phase voltages obtained by SVPWM technique has triple harmonics which is eliminated in line voltage. The peak value of the line voltage is 15% more than that in sine PWM at maximum modulation index, so this method of PWM generation gives better utilization of dc bus voltage for inverter.

CHAPTER 5

DIRECT TORQUE FUZZY CONTROL

5.1 Introduction

Artificial intelligence like neural networks, fuzzy logic and genetic algorithm has a great importance in controlling the variable speed drives. In the previous chapters direct torque control using classical method and space vector pulse width modulation method was discussed. The main drawback of the conventional DTC was high stator flux and torque ripples also the speed of induction motor drives is reduced under transient and dynamic operating condition. The ripples in torque and speed variation in transient and dynamic conditions were compensated using space vector pulse width modulation and fuzzy logic controller based DTC. This chapter mainly focuses on the proposed fuzzy logic control for improving the performance of classical DTC [33].

5.2Proposed fuzzy logic controller

The schematic of the basic functional blocks used to implement the proposed DTFC of induction motor drive is shown in Fig.5.1. A voltage source inverter supplies the motor and instantaneous values of the stator flux and torque are calculated from stator variable by using a closed loop estimator. Stator flux and torque can be controlled directly and independently by properly selecting the inverter switching.

The fuzzy logic controller is characterized as follows:

(a) Seven fuzzy sets for input and output variables,
(b) Fuzzification using continuous universe of discourse,
(c) Implication using Mamdani’s ‘min’ operator,
(d) De-fuzzification using centroid method.

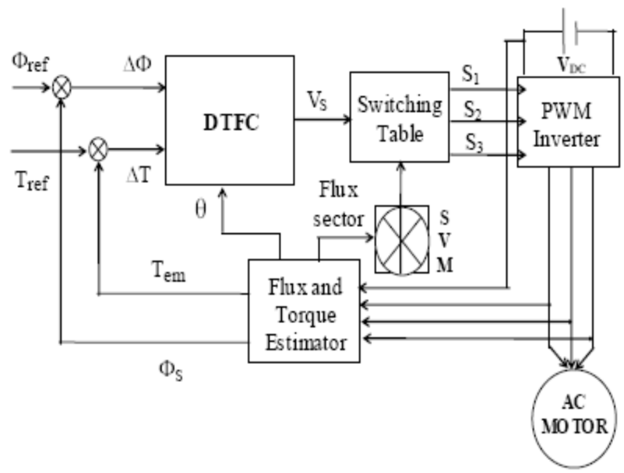

Figure 5.1 Block diagram of proposed FLC for IMD

A fuzzy logic control converts linguistic control strategy into automatic control strategy. The fuzzy rules can be formed using expert knowledge and experience based database. First of all the input torque and the change in torque error are converted into fuzzy variable that can be recognized by level of membership function in fuzzy set. The fuzzy sets are defined with triangular membership functions. In fuzzy membership function there are two input variable and each input variable have seven linguistic values. The fuzzy control rules are shown in table [36,56].

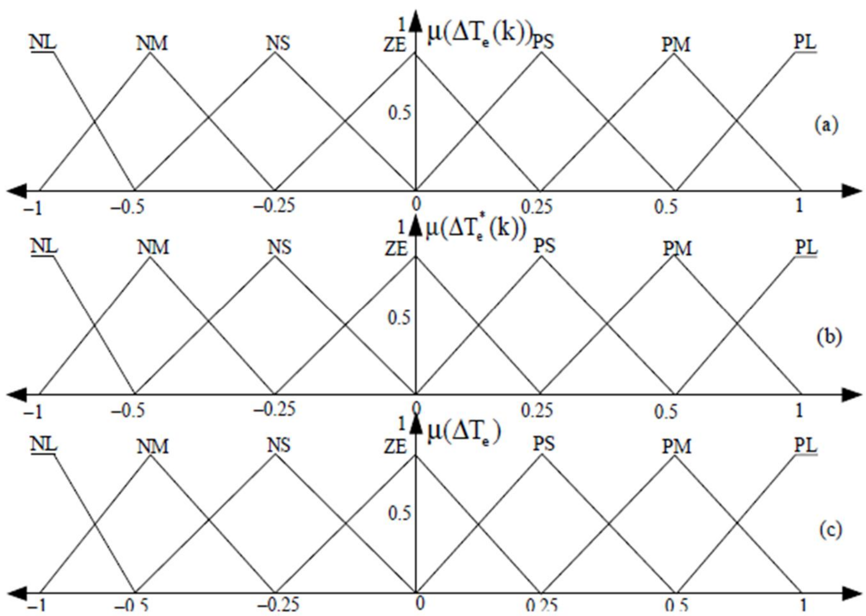

Figure 5.2 the fuzzy membership functions of input variables (a) torque error, (b) change in torque error, and (c) output variable

Table 5.1 Fuzzy logic control rules

$\Delta T_e(k)$ / $\Delta T_e^*(k)$	NL	NM	NS	ZE	PS	PM	PL
NL	NL	NL	NL	NL	NM	NS	ZE
NM	NL	NL	NL	NM	NS	ZE	PS
PS	NL	NL	NM	NS	ZE	PS	PM
ZE	NL	NM	NS	ZE	PS	PM	PL
PS	NM	NS	ZE	PS	PM	PL	PL
PM	NS	ZE	PS	PM	PL	PL	PL
PL	ZE	PS	PM	PL	PL	PL	PL

The simulation result of DTC using fuzzy logic controller is shown in next chapter with discussion of the result.

Surface viewer of change in flux error and torque error is shown below:

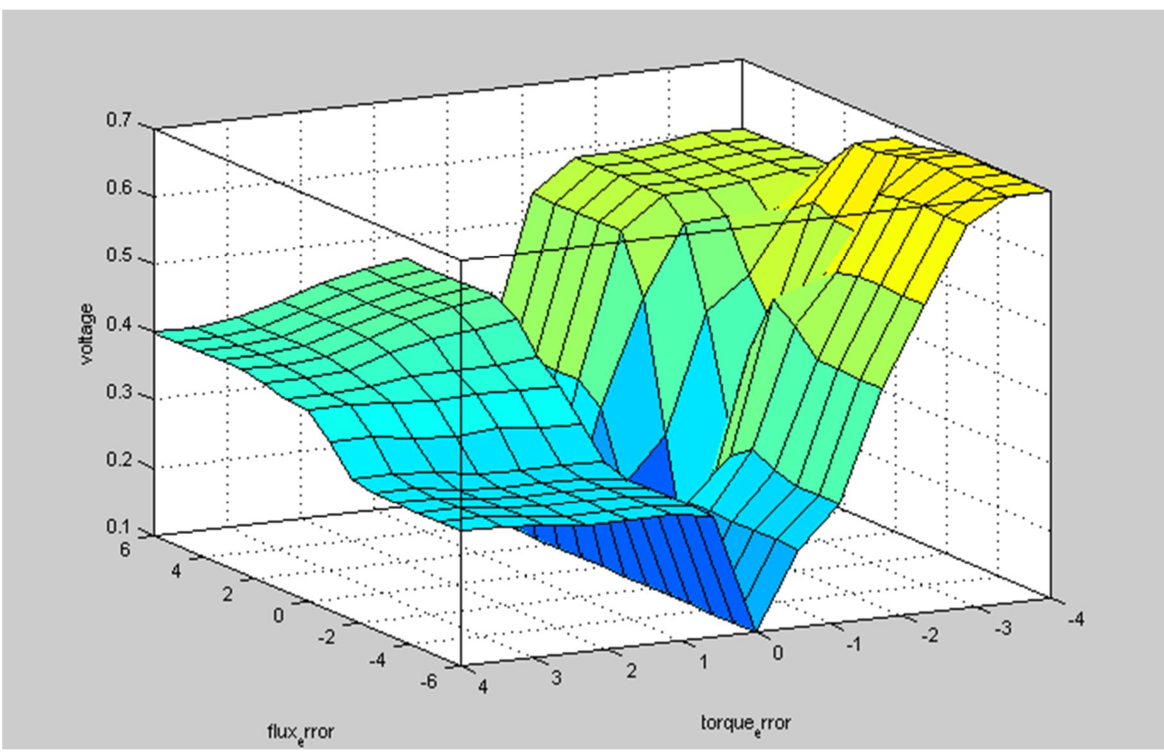

Figure 5.3 surface viewer of fuzzy fis

The rule viewer is also shown:

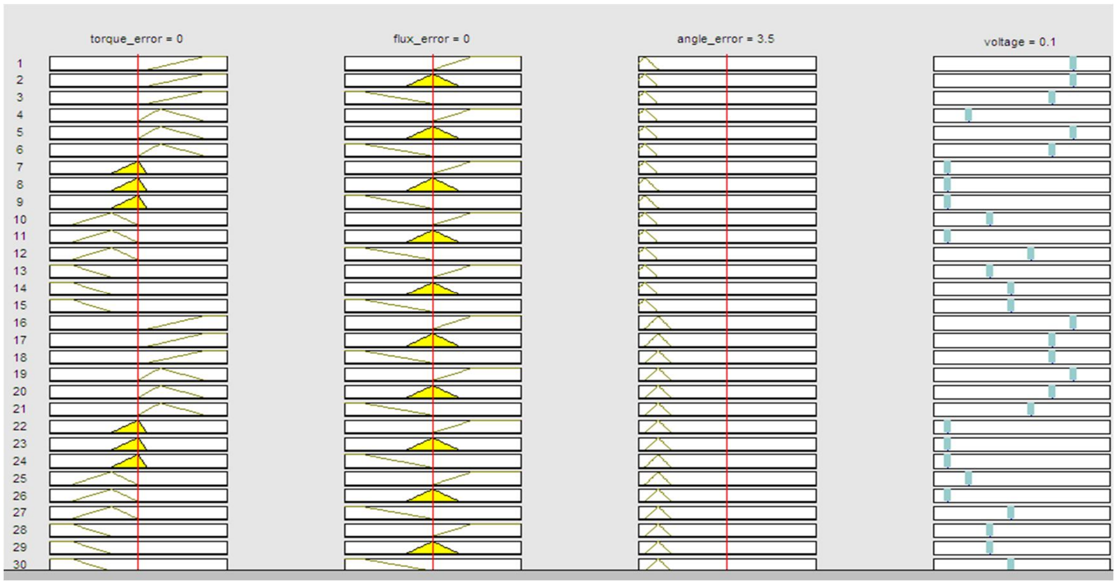

Figure 5.4 rule viewer

5.3 Conclusion:

This chapter deals with the detail view of implementing the fuzzy logic controller in direct torque control of induction motor drive. Triangular membership function is taken and the rule base is formed using the set of linguistic rules. The complete detail of the waveform using fuzzy logic control is shown in the next chapter and is observed that fuzzy logic control gives the better output as compared to the conventional DTC scheme and the ripples in output waveform is also removed. Hence we can conclude that this method is comparable to that of DTC using SVPWM.

CHAPTER 6

SIMULATION MODELS RESULTS AND DISCUSSIONS

6.1 SIMULATION OF DTC SCHEME

A direct torque control algorithm of Induction motor drive has been simulated using MATLAB/Simulink 7.10.0 (R2010a). The induction motor is fed by an IGBT PWM voltage source inverter consists of Universal Bridge Block. Flux and torque references are produced using speed control loop which employs proportional integral control. The DTC block calculates the estimated motor torque and flux and then compares with their references respectively. The comparators outputs are then used by an optimal switching table which generates the inverter switching pulses. The MATLAB/ SIMULINK model for switching logic is developed. The transient performance of the proposed DTC model is tested by applying a load torque command on the mechanical dynamics. The model is run for typical conditions of reference speed and applied torque value. Figure 5.2 depicts the complete Simulink model of DTC scheme of IM. A 200HP, 460V, 60Hz, 4pole, 3-phase induction motor is used for the simulation.

6.2 Description of Simulink model

The speed controller block is based on PI regulator. The output of the regulator is flux and torque set point which is further applied to DTC controller block. The DTC controller block contains five main blocks torque and flux calculator block, $\alpha\beta -$ vector block, flux and torque hysteresis block containing two-label hysteresis comparator for flux control and three-label hysteresis comparator for torque control, switching table block and switching control block. Braking chopper is used to absorb energy produced during motor deceleration. The control system has two sampling times.

- Speed controller sampling time
- DTC controller sampling time.

The sampling time of speed controller should be a multiple of the DTC sampling time.

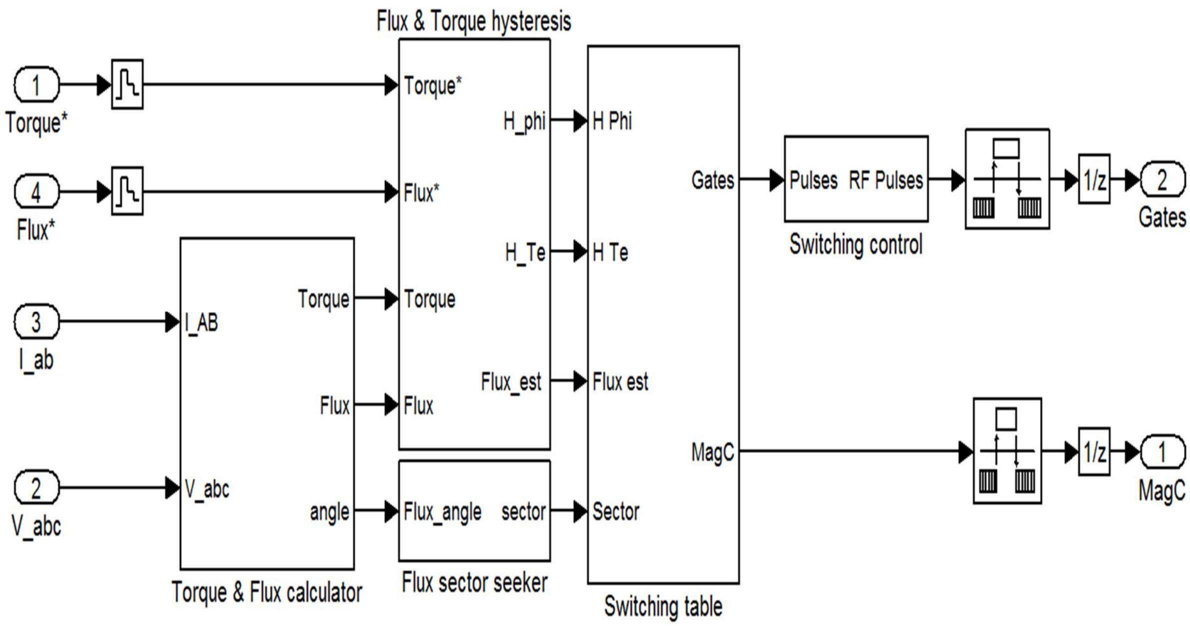

Figure 6.1 Simulink model of DTC controller

6.3 PERFORMANCE EVALUATION OF DTC USING SVPWM

The simulation results obtained are shown in the figure with proposed DTC scheme. This demonstrates that the proposed DTC achieved high dynamic performance in speed response to variations in required torque. However, the performance diminishes with torque overshoot in the torque transient because of the hysteresis controllers used. The figure shows the motor electromagnetic torque and rotor speed. The torque has high initial value in the acceleration zone, increases due to load torque increment then decreases and remains constant in the deceleration zone. The flux also increases in transient period and then it comes to steady state so that flux is maintained constant to control the torque. The q-axis stator flux varies with torque; it has high initial value in the acceleration zone then decreases and comes to steady state. The stator current, rotor speed and electromagnetic torque by using SVPWM method is also shown in figure. We

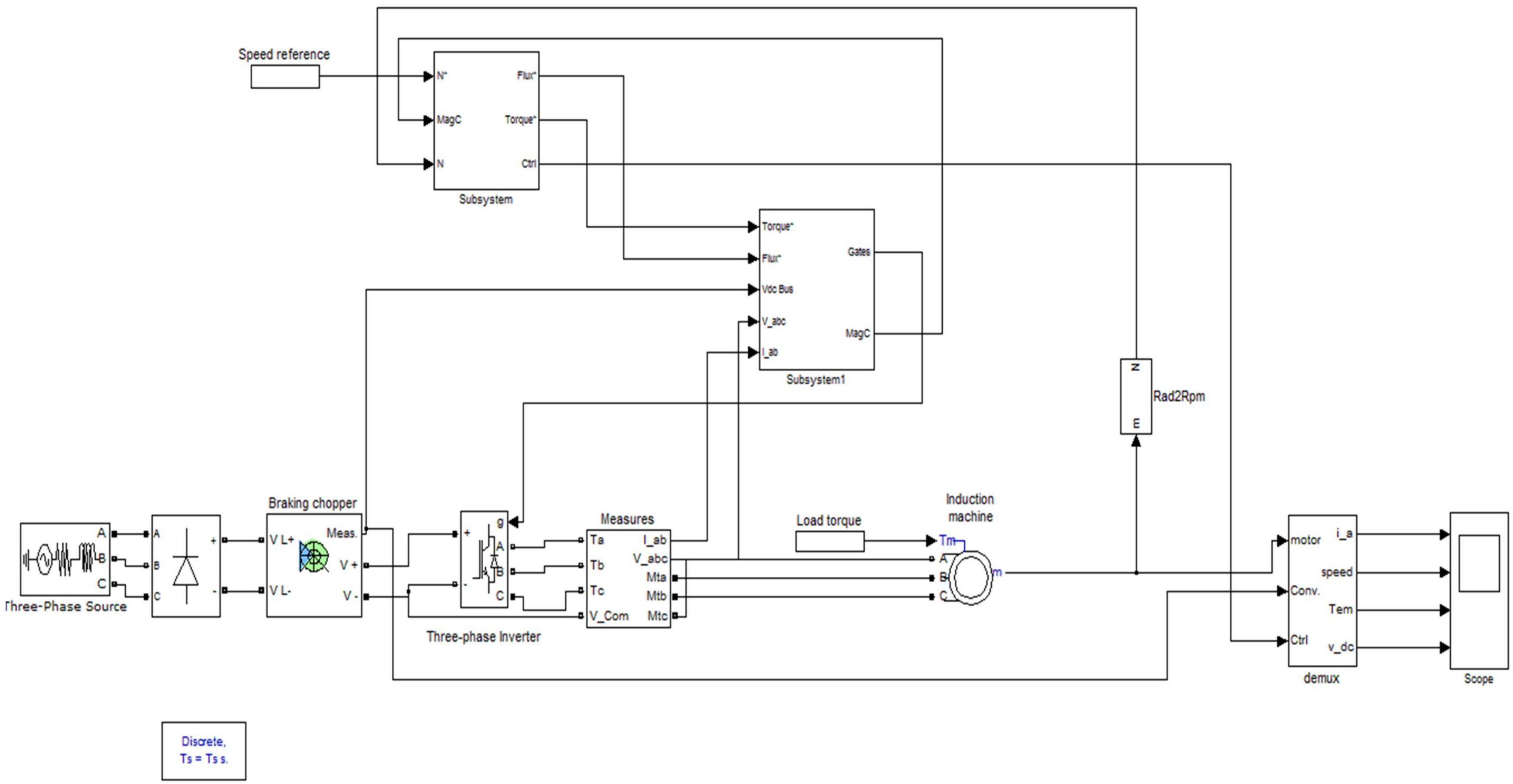

Figure 6.2 Simulink model of DTC Algorithm of IM

can see that the ripple in the waveforms reduces as compared to the conventional DTC scheme and fuzzy logic controller.

Case 1: t = [0,1]; speed = [500,0]

t = [0, 0.2, 0.5, 1.5]; Torque = [0, 200, 792, -792]

Using conventional DTC

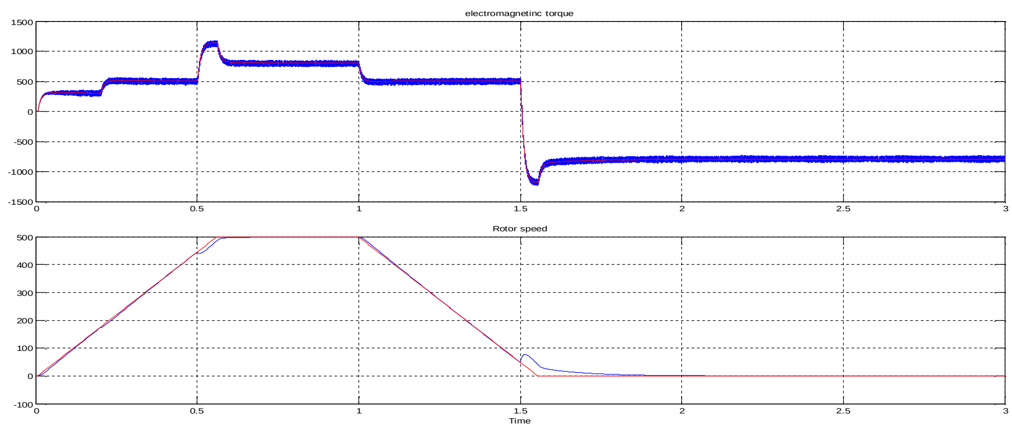

Figure: 6.3(a) Plots of electromagnetic torque and speed using conventional DTC

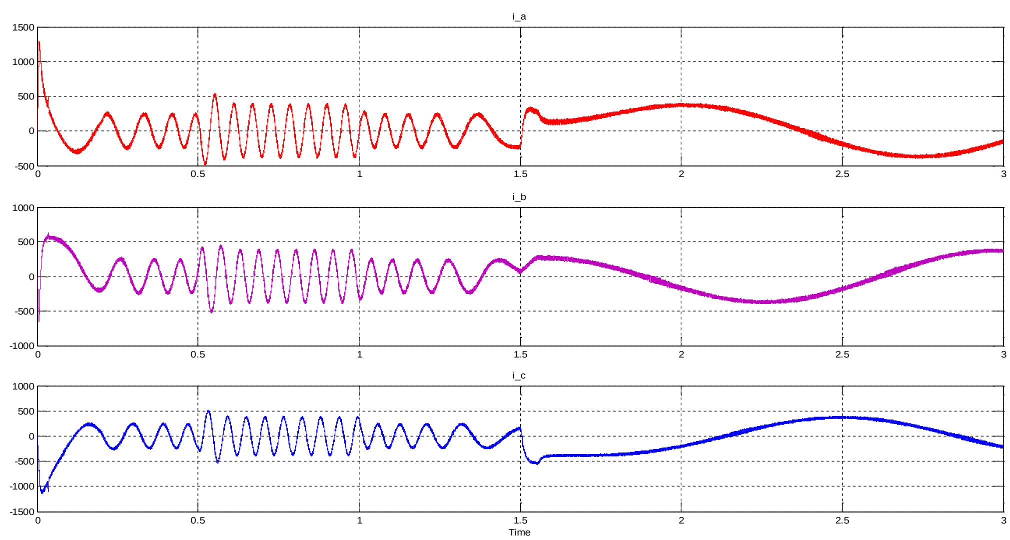

Figure 6.3(b) Plots of stator current using conventional DTC

At t=0 sec speed is set to 500 rpm and torque is set to zero value. Now at t=0.2 sec a load torque of 200 N-m is applied while the motor is still accelerating to its final value. The electromagnetic torque increases to its maximum value at t=0.5 sec and then stabilizes to at 800 N-m. At t=0.5 sec speed set point is changed to 0 rpm, speed decreases to 0 rpm with deceleration and finally stabilizes to zero value. The variation in stator current, electromagnetic torque and rotor speed is shown in fig 6.3(a) and 6.3(b).

t = [0,1]; speed = [500,0]

t = [0, 0.2, 0.5, 1.5]; Torque = [0, 200, 792, -792]

DTC using SVPWM:

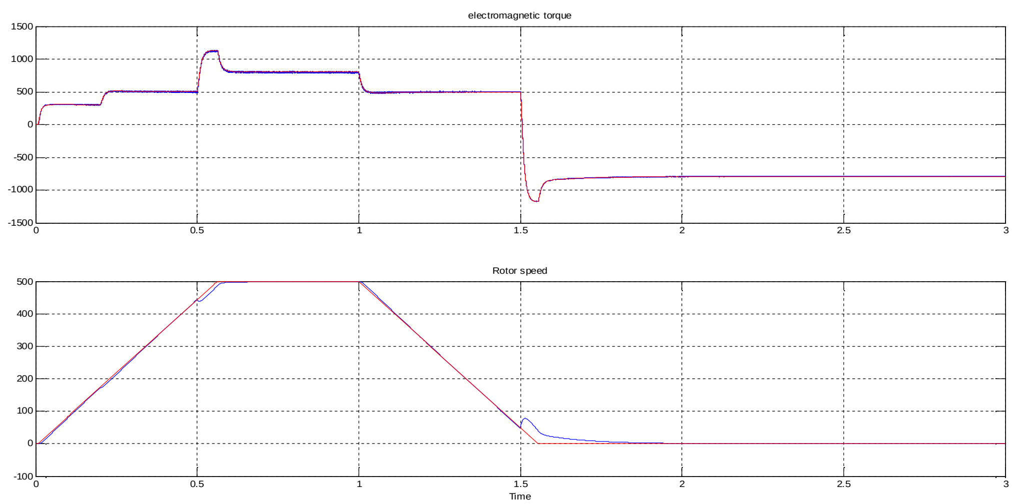

Figure: 6.4(a) Plots of electromagnetic torque and speed using SVPWM DTC

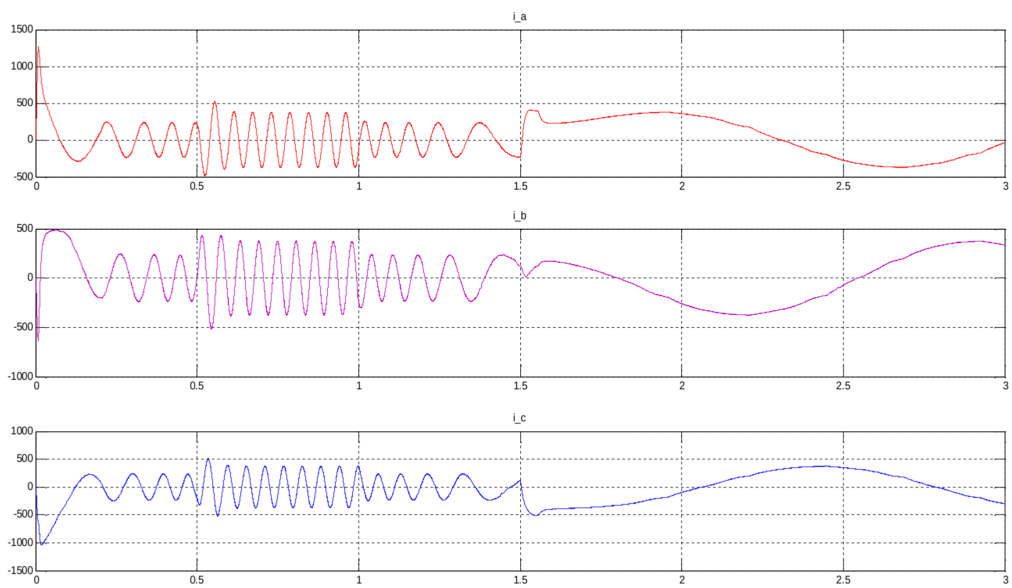

Figure 6.4(b) Plots of stator current using SVPWM DTC

From figure 6.4(a) and 6.4(b) it was observed that the variation in waveform remains same but ripples in output waveforms are substantially reduced. The ripples in electromagnetic torque are removed. At t = 1 sec speed reference is set to zero value and hence the rotor speed is finally stabilizes to zero value following the deceleration. The plots are shown in 6.4(a) and 6.4(b).

t = [0,1]; speed = [500,0]

t = [0, 0.2, 0.5, 1.5]; Torque = [0, 200, 792, -792]

DTC using fuzzy logic controller:

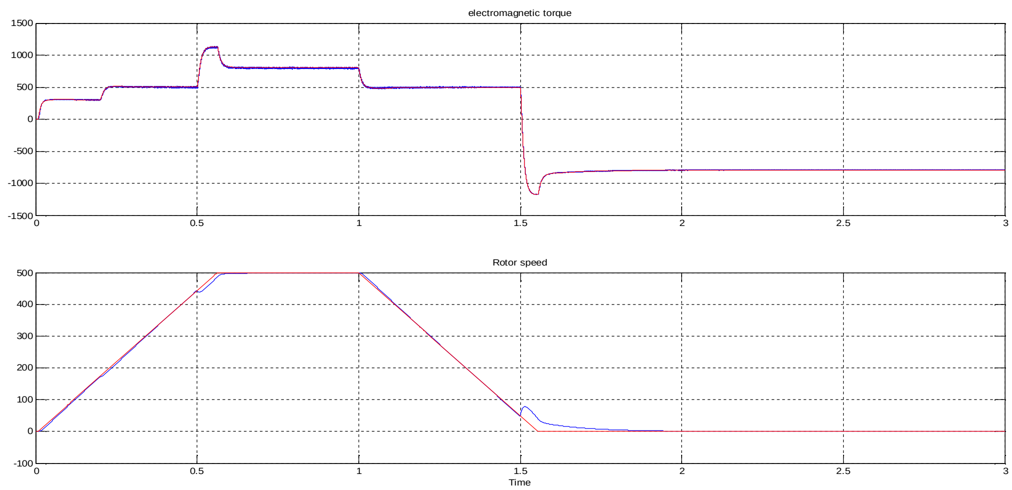

Figure: 6.5(a) Plots of electromagnetic torque and speed using FLC DTC

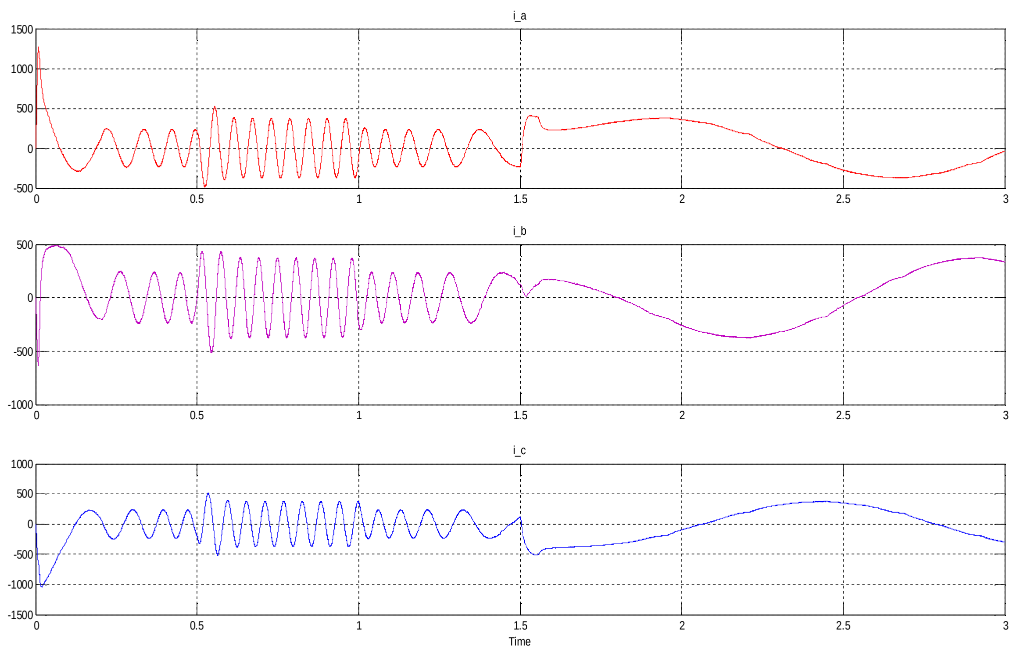

Figure 6.5(b) Plots of stator current using FLC DTC

From the diagram we can conclude that the waveform of the IMD using fuzzy logic control is comparable to the control method using SVPWM. Ripples in the waveform is also reduced which is shown in 6.5(a) and 6.5(b). As the loading condition is changed the variation in stator current, electromagnetic torque and rotor speed is following the SVPWM method and the ripples in waveform is substatially reduced.

Case 2 t = [0, 1]; speed = [500, 100]

t = [0, 0.2, 0.5, 1.5]; torque = [0, 200, 500, 700]

Using conventional DTC:

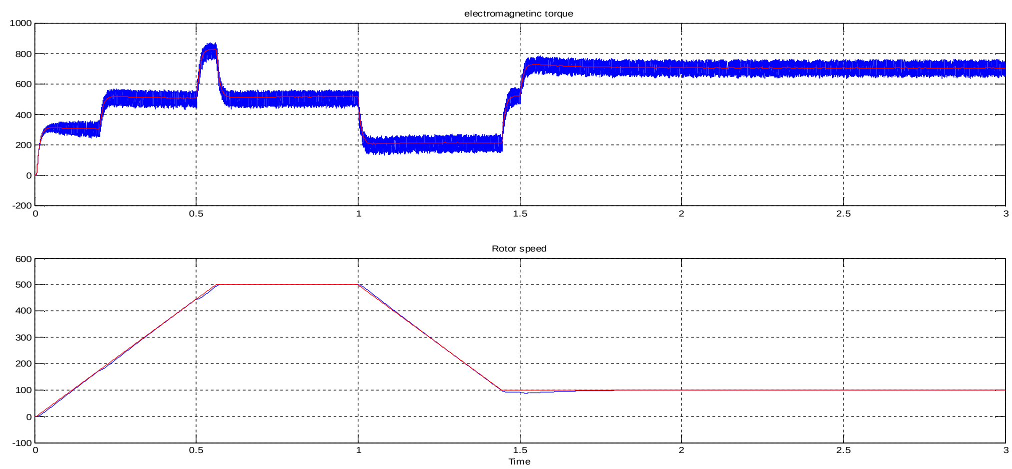

Figure: 6.6(a) Plots of electromagnetic torque and speed using conventional DTC under different loading conditions

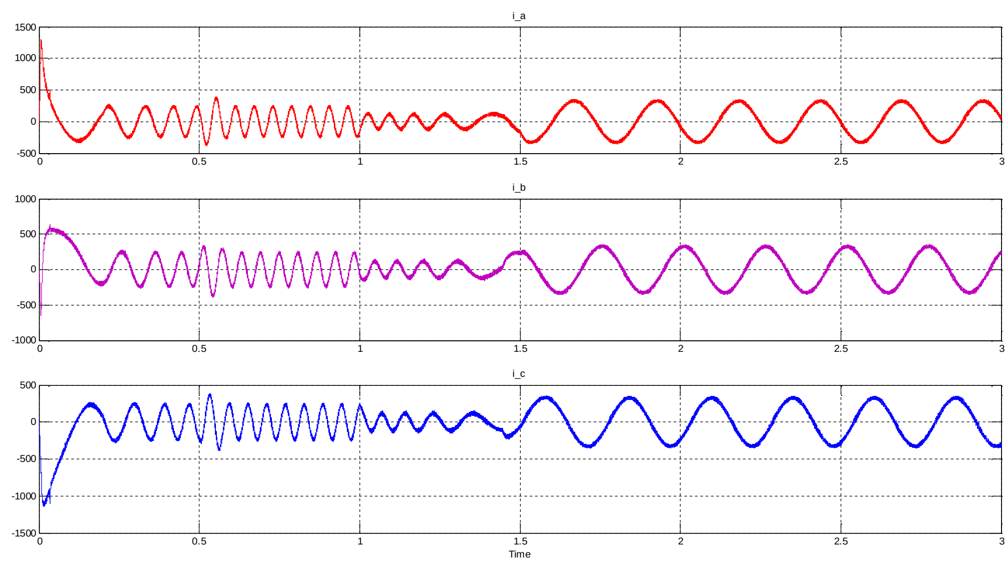

Figure: 6.6(b) Plots of stator current using conventional DTC under different loading conditions

At t=0 sec speed is set to 500 rpm and torque is set to zero value. Now at t=0.2 sec a load torque of 200 N-m is applied while the motor is accelerating and at t = 0.5 sec a load af 500 N-m is applied due to which the speed is still accelerating and stabilizes to its final value. The electromagnetic torque increases to its maximum value at t=0.55 sec and then again at t = 1.5 sec load torque is changed to 700 N-m hence the final value of load torque is stabilizes to 700 N-m. At t = 1 sec speed set point is changed to 100 rpm, and hence finally speed value sets at 100 rpm. The variation in stator current, electromagnetic torque and rotor speed is shown in fig 6.6(a) and 6.6(b).

t = [0, 1]; speed = [500, 100]

t = [0, 0.2, 0.5, 1.5]; torque = [0, 200, 500, 700]

DTC using SVPWM:

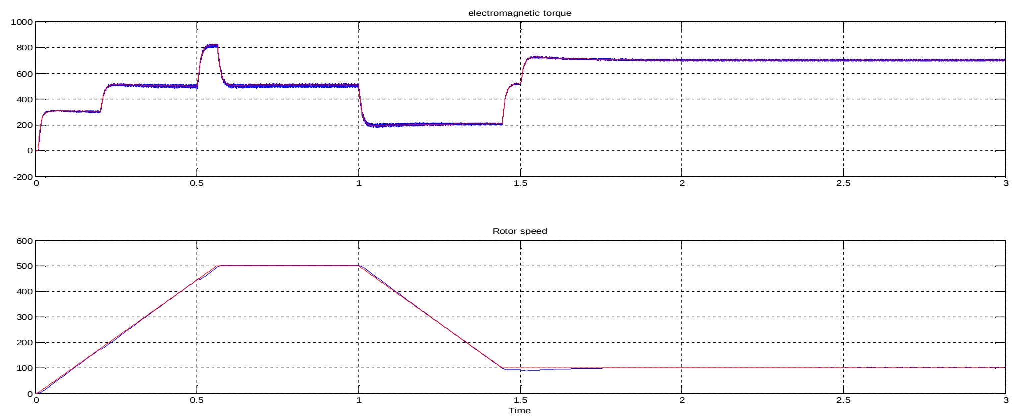

Figure: 6.7(a) Plots electromagnetic torque and speed using SVPWM under different loading conditions

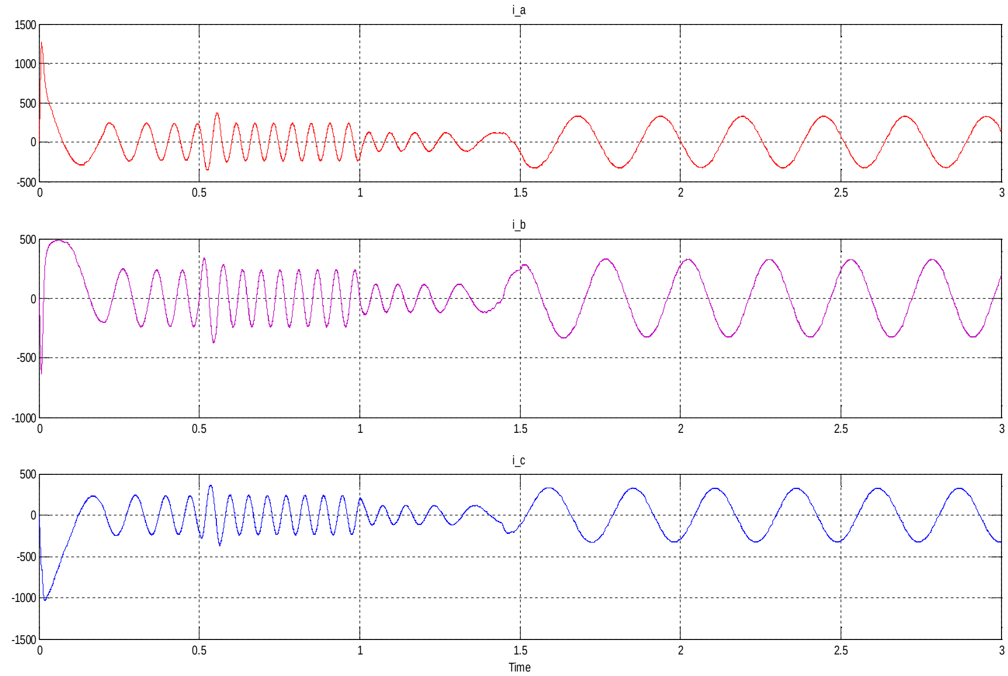

Figure: 6.7(b) Plots stator currents using SVPWM under different loading conditions

From figure 6.7(a) and 6.7(b) it was observed that the variation in waveform remains same but ripples in output waveforms are substantially reduced. The ripples in electromagnetic torque are removed. At t = 1 sec speed reference is set to 100 value and hence the rotor speed is finally stabilizes to 100 value following the deceleration.

t = [0, 1]; speed = [500, 100]

t = [0, 0.2, 0.5, 1.5]; torque = [0, 200, 500, 700]

DTC using fuzzy control:

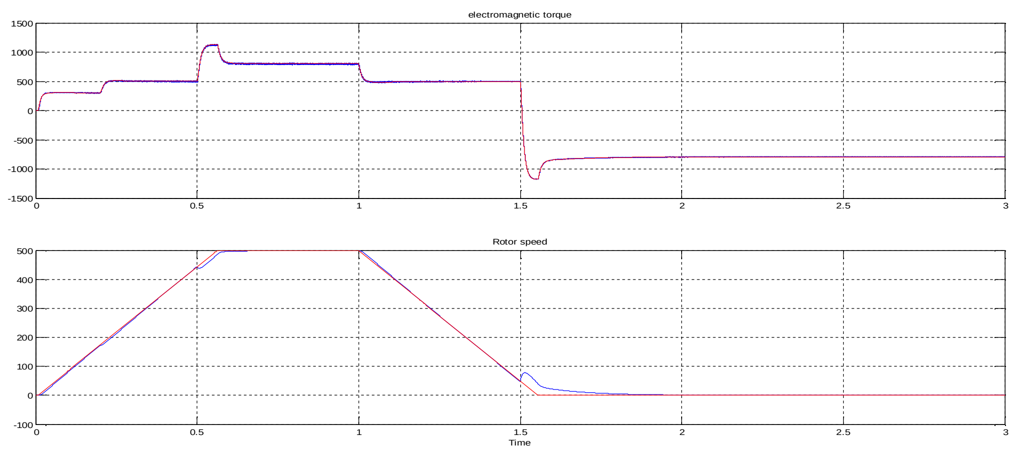

Figure 6.8(a) plots of electromagnetic torque and speed using Fuzzy logic controller

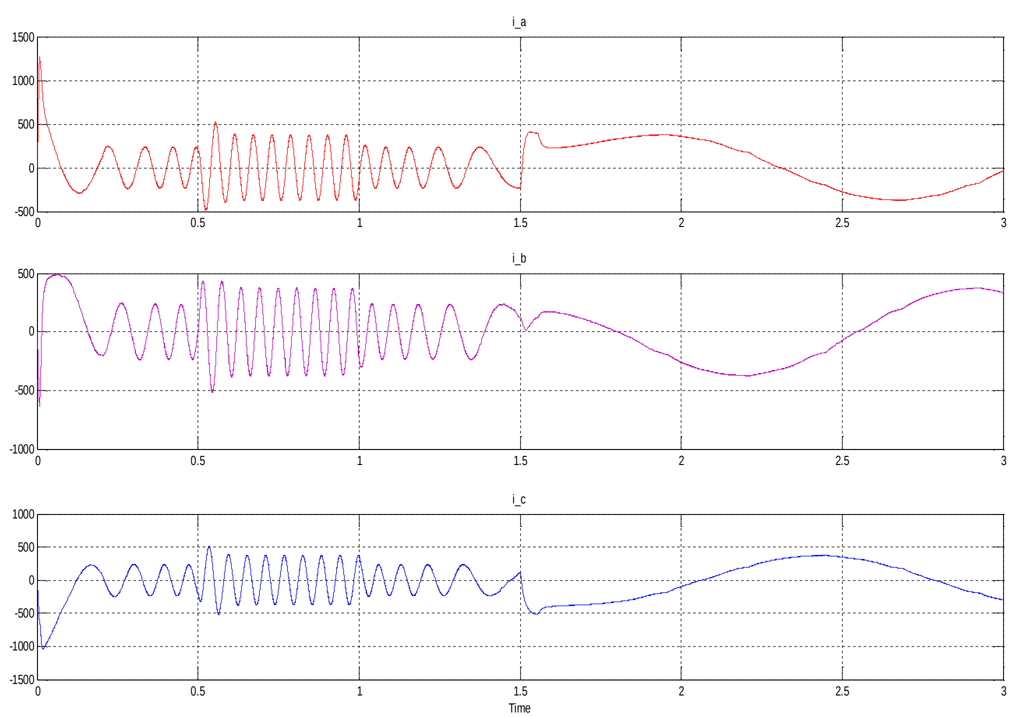

Figure 6.8(b) Plot of stator current of DTC using fuzzy logic controller

From the diagram we can conclude that the waveform of the IMD using fuzzy logic control is comparable to the control method using SVPWM. Ripples in the waveform is also reduced which is shown in 6.8(a) and 6.8(b).

Case 3:

t = [0]; speed = [500]

t = [0, 0.5]; Torque = [0, 500]

Using conventional DTC:

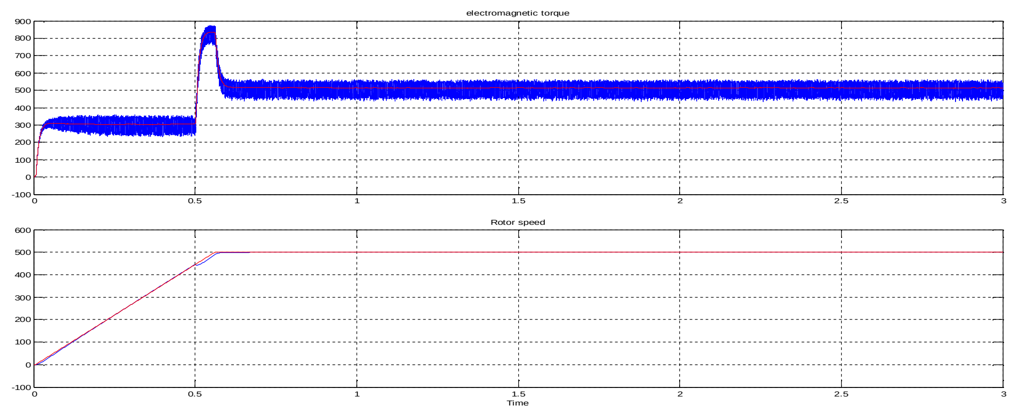

Figure 6.9(a) Plots of torque and rotor speed for a constant speed using conventional DTC

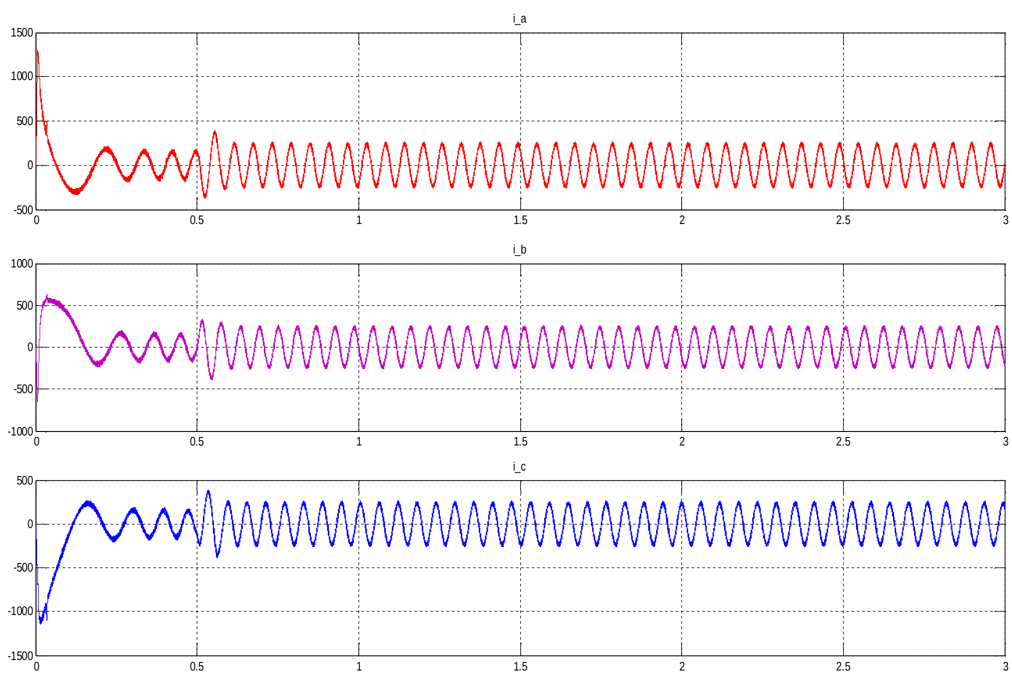

Figure 6.9 (b) Plots of stator current for a constatnt speed using conventional DTC

t = [0]; speed = [500]

t = [0, 0.5]; Torque = [0, 500]

DTC using SVPWM:

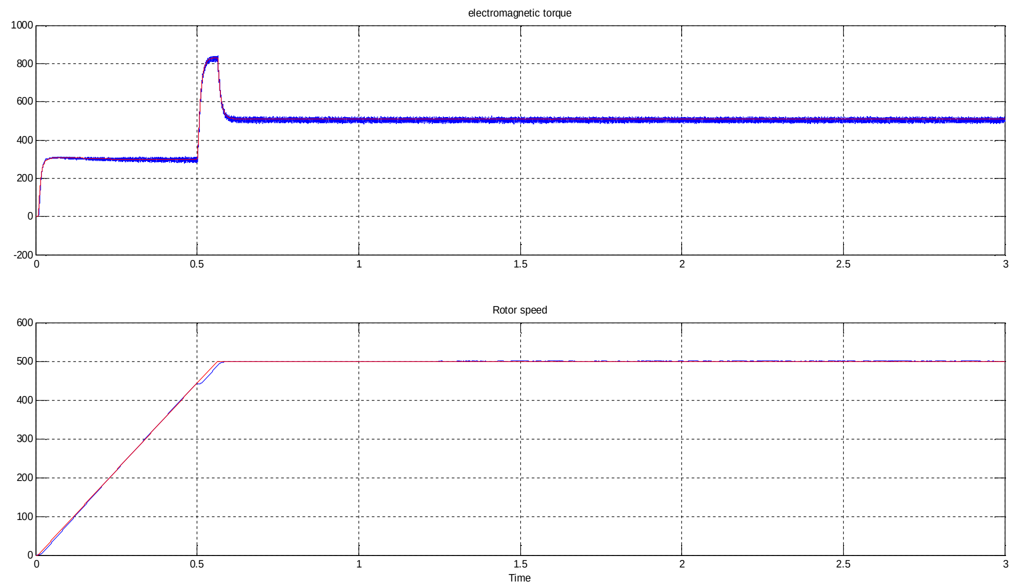

Figure 6.10 (a) Plots of torque and speed for a constatnt speed using SVPWM DTC

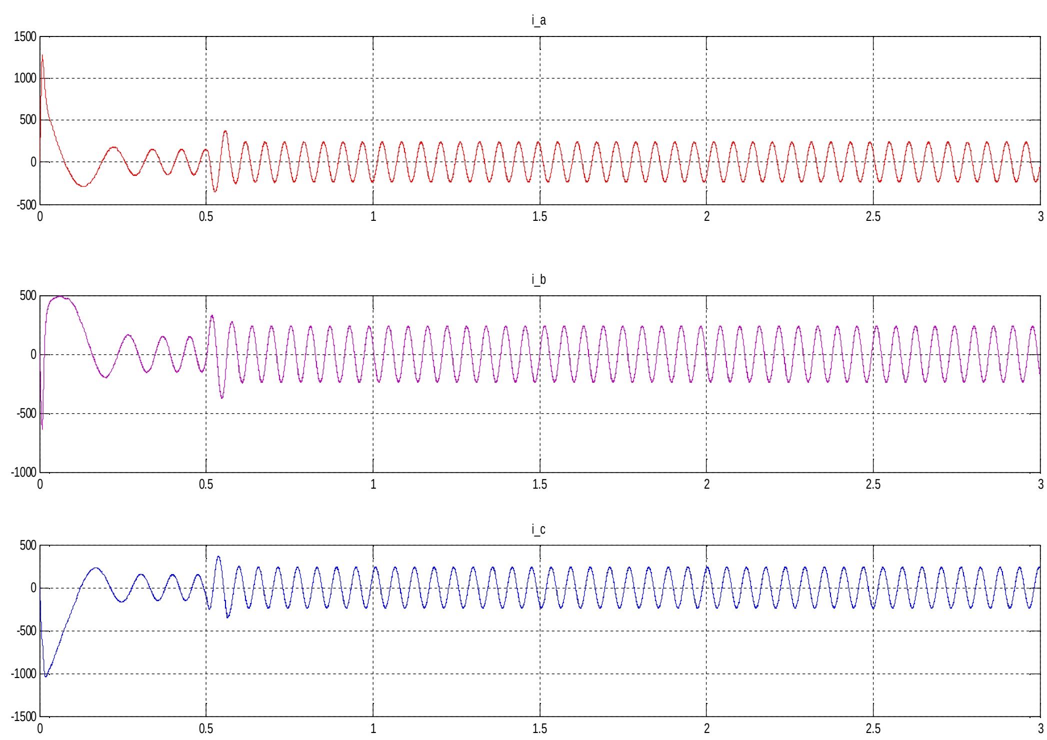

Figure 6.10(b) Plots stator current for a constatnt speed using SVPWM DTC

t = [0]; speed = [500]

t = [0, 0.5]; Torque = [0, 500]

DTC using FLC:

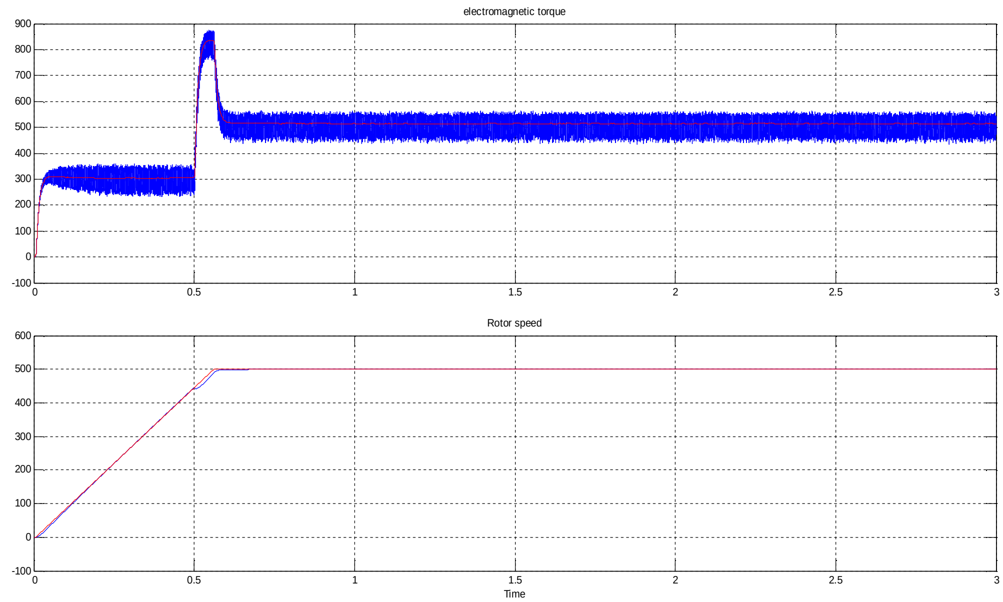

Figure 6.11 (a) Plots of torque and speed for a constatnt speed using FLC DTC

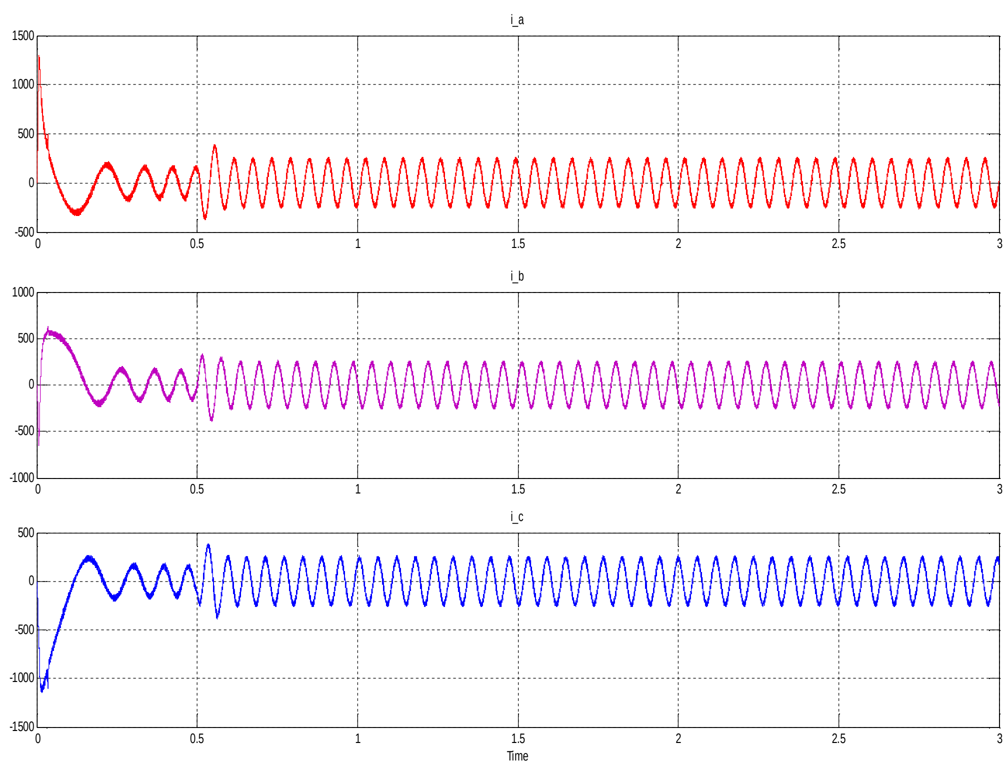

Figure 6.11(b) Plots stator current for a constatnt speed using FLC DTC

From the diagram we can conclude that the waveform of the IMD using conventional DTC, SVPWM and fuzzy logic control that at t = 0.5 sec a load of 500 N-m is applieddue to which stator current fluctuates for a cycle and then comes to a steady state value. After then the motor runs with a constant speed of 500 rpm. The variation in stator current, electromagnetic torque and rotor speed is shown in the figure 6.9 to 6.11.

CHAPTER 7

CONCLUSION AND FUTURE SCOPE OF WORK

7.1 CONCLUSION

For any IM drives, direct torque control is one of the best controllers proposed so far. It allows decoupled control of motor stator flux and electromagnetic torque. From the analysis it is proved that, this strategy of IM control is simpler to implement than other vector control methods as it does not require pulse width modulator and co-ordinate transformations. But it introduces undesired torque and current ripple. DTC scheme uses stationary *d*-*q* reference frame with *d*-axis aligned with the stator axis. Stator voltage space vector defined in this reference frame control the torque and flux. The main inferences from this work are:

- In transient state, by selecting the fastest accelerating voltage vector which produces Maximum slip frequency, highest torque response can be obtained.
- In steady state, the torque can be maintained constant with small switching frequency by the torque hysteresis comparator by selecting the accelerating vector and the zero voltage vectors alternately.
- In order to get the optimum efficiency in steady state and the highest torque response in transient state at the same time, the flux level can be automatically adjusted.
- If the switching frequency is extremely low, the control circuit makes some drift which can be compensated easily to minimize the machine parameter variation.

The estimation accuracy of stator flux is very much essential which mostly depends on stator resistance because an error in stator flux estimation will affect the behavior of both torque and flux control loops. The torque and current ripple is minimized by employing space vector modulation technique.

7.2 FUTURE SCOPE

The conventional DTC scheme using space vector modulation technique was done in the thesis work which can further be improved using hybrid space vector PWM technique which will be helpful in reducing the switching loss and the steady state ripple in torque, flux and current. This method can be used for controlling electric drives of small and medium ratings. Double three phase inverter can also be implemented to smooth the torque of the machine. The torque ripple in DTC scheme can also be improved using discrete pulse width modulation technique. The ripples in output torque and excess of stator current can be removed using neural network or ANFIS and also by using metahuristic algorithms.

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APPENDIX

SYSTEM DATA

THREE PHASE INDUCTION MOTOR RATING:

- 200 hp / 460 Volts/ 60 HZ.
- Stator resistance per phase 14.85×10^{-3} ohm
- Stator leakage inductance 0.303×10^{-3} H
- Rotor resistance per phase 9.295×10^{-3} ohm
- Rotor leakage inductance 0.303×10^{-3} H
- Mutual inductance 10.46×10^{-3} H
- No. of poles 4
- Rotor inertia 3.1Kg-m^2

INVERTER:

- Device type – IGBT/ Diodes

PI REGULATOR

- k_p= 30
- k_i = 200
- DTC sampling time 20 microseconds.