COMPARATIVE EVALUATION OF VECTOR CONTROL AND DIRECT TORQUE CONTROL INDUCTION MOTOR DRIVES

A DISSERTATION

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(With Specialization in Power Apparatus & Electric Drives)

By

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CANDIDATE'S DECLARATION

I hereby declare that the work, which is presented in this dissertation report, entitled **"COMPARATIVE EVALUATION OF VECTOR CONTROL AND DIRECT TORQUE CONTROL INDUCTION MOTOR DRIVES"**, being submitted in partial fulfillment of the requirements for the award of the degree of **MASTER OF TECHNOLOGY** with specialization in **POWER APPARATUS AND ELECTRIC DRIVES**, in the Department of Electrical Engineering, Indian Institute of Technology, Roorkee, is an authentic record of my own work carried out from July 2007 to June 2008, under the guidance and supervision of **Dr. Pramod Agarwal**, Professor, Department of Electrical Engineering and **Dr. S. Ghatak Choudhuri**, Asst. Professor, Department of Electrical Engineering, Indian Institute of Technology, Roorkee.

The results embodied in this dissertation have not submitted for the award of any other Degree or Diploma.

Date: 30/06/2008
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CERTIFICATE

This is to certify that the statement made by the candidate is correct to the best of my knowledge and belief.

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ABSTRACT

Vector Control and Direct Torque Control are becoming the industrial standards for induction motor control. The vector control technique decouples the two components of stator current: one providing the control of flux and the other providing the control of torque. The flux as well as the torque level of the machine is controlled with perpendicular components of the stator current vector in the synchronously rotating reference frame (SRRF). Thus a current control loop is usually realized which controls the stator current and calculates the necessary inverter switching states.

In Direct torque control to achieve the decoupling of flux and torque components, the flux and the torque errors are processed in hysteresis controller. The inverter switching states necessary to achieve the flux and torque levels are then directly obtained from a switching table with the stator flux angle as additional input. The motor torque and flux become direct controlled variables and hence, the name – Direct Torque Control.

This work gives a contribution for a detailed comparison between the two control techniques, emphasizing upon the advantages and disadvantages respectively. The performance of the two control schemes is evaluated based on various criteria including the complexity of control and the dynamic performance. The dynamic response of vector control and direct torque control under various operating conditions such as starting, speed reversal and load perturbation (load application and load removal) is simulated and examined in MATLAB 7.3 environment using simulink and power system blockset toolboxes. The analysis has been carried out on the basis of the results obtained by numerical simulations. The simulation and evaluation of both control strategies are performed using voltage source inverter fed three phase squirrel cage induction motor for 1HP and 30HP ratings respectively.

CONTENTS

List of Symbols

v_{as}, v_{bs}, v_{cs}	stator voltages
i_{as}, i_{bs}, i_{cs}	stator currents
i_{ds}, i_{qs}	direct and quadrature axis stator currents
$\overline{V_s}, \overline{I_s}$	stator voltage and current space vectors
ω_m , ω_r	rotor speed(mech rad/sec, elec rad/sec resp)
V_{dc}	dc link voltage
S_a, S_b, S_c	inverter gating signals(0 or 1)
$\overline{\psi_s}, \overline{\psi_r}$	stator and rotor flux linkages
L_{ls}, L_{lr}	stator and rotor leakage inductances
L_s, L_r, L_m	stator, rotor and mutual inductances
R_s, R_r	stator and rotor resistances
$\tau_r = L_r / R_r$	rotor time constant
T_e, T_l	electromagnetic and load torque resp
$\sigma = 1 - L_m^2 / L_s L_r$	total leakage factor
s	laplace operator
p	number of poles
$e^{j\psi}$	vector rotator
$i_{as}^*, i_{bs}^*, i_{cs}^*$	reference stator currents
T_e^*	reference electromagnetic torque

i_{mr}	magnetizing current
J	moment of inertia
K	torque proportionality constant
K_p	proportional gain constant of speed controller
K_i	integral gain constant of speed controller
M	mutual inductance
σ_r	rotor leakage factor
ω_2	slip speed
ω_{mr}	synchronous speed of the motor
γ	angle between the stator and rotor flux linkage space vectors
α	angle between the stator flux linkage and stator current space vector
$\overline{\lvert\psi_{sref}\rvert}$	reference value of the stator flux-linkage space vector
$\overline{\lvert\psi_s\rvert}$	actual modulus of the stator flux-linkage space vector

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CHAPTER I

INTRODUCTION

1.0 GENERAL

Faraday discovered electro-magnetic induction in 1831. After this discovery, it took about fifty years for the development of the first induction machine by Nikola Tesla in 1888. The development of this electrical machine that do not require brushes for its operation marked a revolution in electrical engineering and gave a scope for the wide spread use of poly-phase generation and distribution systems. The choice of present mains frequency (60 Hz in the USA and 50 Hz in Europe and Asia) was established in the late. 19^{th} century because Tesla found it suitable for his induction motors, 60 Hz was found to produce no flickering when used for lightening applications.

Now-a-days, more than 60% of all the electrical energy generated in the world is used by cage induction motors. Induction machines have been mostly used at fixed speed for more than a century. On the other hand, DC machines have been used for variable speed applications

In DC machines armature mmf axis is established at 90° electrical to the main field axis. The electromagnetic torque is proportional to the product of field flux and armature current [1]. Field flux is directly proportional to the field current and is unaffected by the armature current because of orthogonal orientation between armature mmf and field mmf. Therefore in a separately excited DC machine, with a constant value of field flux, torque is directly proportional to armature current. Hence direct control of armature current gives direct control of motor torque and fast response; because motor torque can be altered as rapidly as armature current can be altered. Hence dc motors are simple in control and offer better dynamic response inherently. Numerous economical reasons, for instance high initial cost, high maintenance cost for commutators, brushes and brush holders of dc motors call for a substitute which is capable of eliminating the persisting problems in dc motors and has all the advantages present in dc motors. Freedom from regular maintenance and a brushless robust structure of the three phase squirrel cage induction motor are among the prime reasons, which brings the motor forward as a good substitute.

The ac induction motors are the most common motors used in industrial motion control systems, as well as in main powered home appliances [5]. Simple and rugged design, low-cost and low maintenance are some of the main advantages of 3 phase ac induction motors. Various types of 3 phase ac induction motors are available in the market. Different motors are suitable for different applications. The speed and the torque control in case of 3 phase ac induction motors require a greater understanding of the design and the characteristics of these motors.

1.1 CONTROL TECHNIQUES OF INDUCTION MOTOR

Before the advent of vector control techniques [2,3] for induction motors, various control methods ex: voltage control, frequency control, rotor resistance control, v/f control, flux control [2, 4] etc., have been used and named as scalar control. . In scalar control, the motor is fed with variable frequency supply generated by an inverter controlled by Pulse Width Modulation (PWM) techniques. Here, the V/f ratio is maintained constant in order to get constant torque through out the operating range of the motor. Since only magnitudes of the input variables – frequency and voltage – are controlled, this is known as "scalar control" such control techniques result in satisfactory response only in steady state conditions. The dynamic response of the drive from scalar control methods is observed to be sluggish. This is one of the major drawbacks of scalar control technique as applied to the induction motors. It had resulted in development of the variable voltage-variable frequency (VVVF) control technique, which could provide improved response of an induction motor drive in transient as well as in steady state operating conditions.

In general, the drives with this type of control are without any feedback devices i.e. open loop control. Hence, a control of this type offers low cost, easy to implement and also a very little knowledge of the motor is required for frequency control. Thus, this type of control is widely used. But there are certain disadvantages with this type of control which includes, the torque developed is load dependent and is not controlled directly. Also, the transient response of such a control is not fast due to the predefined switching pattern of the inverter. Now variable speed applications by induction motors came into existence with the arrival of solid-state power electronic devices like thyristors, IGBTS and MOSFETS, etc. These provide high performance torque, speed and flux control. Although induction machines using V/f control can maintain a constant flux, their torque and flux dynamic performance is extremely poor. The advantages of induction machines are clear in terms of robustness and price; however it was not until the development and implementation of Vector Control (VC) or Field Oriented Control (FOC) that induction machines were able to compete with DC machines in high performance applications.

The principle behind FOC is that the machine flux and torque are controlled independently, in a similar fashion to a separately excited DC machine. Instantaneous stator currents are "transformed" to a reference frame rotating at synchronous speed aligned with the rotor, stator or air-gap flux vectors, to produce a d-axis component current(flux producing) and a q-axis component of current(torque producing)[4]. Such a rotating frame of reference is referred as the synchronously rotating reference frame (SRRF). In this work, SRRF is aligned with rotor mmf space vector. In the SRRF, the stator current space vector is split into two decoupled components, one controls flux and the other controls the torque respectively.

An induction motor is said to be in vector control mode, if the decoupled components of the stator current space vector and the reference decoupled components defined by the vector controller in the SRRF match each other respectively. Alternately, instead of matching the two-phase currents (reference and actual) in the SRRF, the close match can also be made in three phase currents (reference and actual) in the stationary reference frame. Hence in spite of induction machine's nonlinear and highly interacting multivariable control structure, its control has become easy with the help of FOC. Therefore FOC technique operates the induction motor like a separately excited DC motor.

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 Stator flux oriented control, Rotor flux oriented control and 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. Depending on the method of measurement, the vector control is divided into two subcategories: direct and 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 measured from the rotor speed, the stator current and the voltage.

1.2 COMPARISON OF BASIC CONTROL CHARACTERISTICS

Both FOC and DTC schemes aim at controlling the motor torque and flux over a speed range. In this section, the basic control characteristics of FOC and DTC are studied and compared with a view to defining the nature of the control strategies and examining their influence on the drive performance and on the implementation complexity [26]. Because of the efficient, smooth and maintenance free operation of VCIMDs, such drives are finding increasing applications in many drive applications such as air-conditioning, refrigerators, fans, blowers, pumps, waste water treatment plants, elevators, lifts, traction, electric vehicles etc.

Flux and Torque Control Mechanism:

FOC scheme uses a d-q coordinates frame having the d-axis aligned with rotor flux vector that rotates at the stator frequency. This particular situation allows the flux and torque to be separately controlled by the stator current d-q components. The rotor flux is a flux of the d-axis component stator current i_{ds}. The developed torque is controlled by the q-axis component of stator current i_{qs}. The decoupling between flux and torque control is achieved only if the rotor flux position is accurately known. This can be done by using direct flux sensors or by using a flux estimator that computes the rotor flux vector from stator voltages and currents and/or speed sensor signal.

On the other hand, direct torque control scheme uses a stationary d-q reference frame having d-axis aligned with stator a-axis. Torque and flux are controlled by the stator voltage space vector $\overline{V_s}$ defined in this reference frame. The variation of the stator flux space vector due to the application of the stator voltage vector $\overline{V_s}$ during a time interval of Δt can be approximated as:

$$\overline{\Delta\psi_s} = \overline{V_s}.\Delta t$$

If the stator resistance R_s is neglected.

Since the rotor flux changes slowly, the rapid variation of stator flux space vector will produce a variation in the developed torque because of the variation of the angle between the two vectors.

Estimated variables:

The operation of both FOC and DTC depends on the system variables that are computed or estimated from the measured quantities [26]. The accuracy of the estimated variables has a direct influence on the control performance.

In FOC system, the estimated variable is the rotor flux angle required for coordinate transformation. The calculation of ψ requires the measured rotor speed ω_r and estimated slip frequency, the slip frequency depends on the time constant and the estimated rotor flux amplitude. The electrical angle is of first importance in FOC operation.

In DTC system, the estimated quantities are the stator flux and motor torque which are required for feedback control. The stator flux is calculated from stator voltage and current space vectors and the developed torque is calculated from the stator flux and the stator current space vectors. The accuracy of stator flux depends mostly on the estimation accuracy on the stator resistance R_s An error on the stator flux will affect the behavior of both flux and torque control loops.

1.3 LITERATURE REVIEW

In 1972, F. Blaschke," The Principle of Field Orientation as applied to the New Transvector Closed-loop Control System for Rotating-Field machines" described the principle of field orientation- a new closed loop control method for rotating field machines. It is shown how these manipulated variables must be influenced to provide instantaneous and well damped adjustment of the torque independently of the inherent characteristics of an induction motor.

In 1986, I. Takahashi, N. Noguchi, "A new quick response and high efficiency control strategy of an induction motor", the proposed scheme is based on limit cycle control of both flux and torque using optimum PWM output voltage, a switching table is employed for selecting the optimum inverter output voltage vectors so as to attain as fast a torque response. The efficiency optimization in the steady state operation is also considered; it can be achieved by controlling the amplitude of the flux in accordance with the torque command.

In 1988, M. Depenbrock, "Direct self-control (DSC) of inverter-fed induction machine", proposed a method to control the torque of an induction motor, it is sufficient to process the measured signals of the stator currents and the total flux linkages only. Optimal performance of drive systems is accomplished in steady state as well as under transient conditions by combination of several two limit controls.

In 1994, D. Casadei, G. Grandi, G. Serra and A. Tani, Effects of flux and torque hysteresis band amplitude in direct torque control of induction machines [18]. In this paper the influence of the amplitude of flux and torque hysteresis bands on switching frequency, torque and flux ripple, current distortion and drive losses is investigated. The advantages of a careful choice of hysteresis band amplitude are emphasized. The results obtained show that the flux hysteresis band mainly affects the motor current distortion in terms of low order harmonics.

In 1995, I. Ludtke and M. G. Jayne investigated the performance of different three phase cage rotor induction motor control systems [24]. Field oriented methods namely direct and indirect vector controls are investigated in both voltage and current controlled forms. Further the possibility of field orientation aligned with the stator, air gap and rotor flux vector is considered. Direct torque control methods are examined and found that DTC is less complex and gives better control characteristics than the vector control methods.

In 2000, D. Telford, M. W. Dunnigan and B. W. Williams provided [29] a comprehensive comparison of the two main high performance induction machine torque control methods ., vector control and direct torque control. It has been shown that both techniques achieve a similar level of transient torque performance, however the DTC scheme is disadvantaged due to the possibility of loss of flux control at lower speeds/loads, higher torque /current ripple and the uncontrolled current transients. The vector control scheme however is disadvantaged by greater parameter sensitivity and the commissioning problems associated with setting up the current control loops.

In 2001, Domenico Casadei, G. Serra et al., presented a new control method[31] for matrix converters which allows, under the constraint of unity input power factor, the generation of the voltage vectors required to implement the direct torque control of induction machines. The current and the torque waveforms emphasize the effectiveness of the control scheme.

Mirjana Milosevic proposed "Hysteresis current control in three phase voltage source inverter" this paper provides different current control methods which play an important role in power electronic circuits, particularly in current regulated PWM inverters which are widely applied in ac motor drives and continuous ac power supplies where the objective is to produce a sinusoidal ac output.

In 2002, FOC and DTC: Two viable schemes for induction motor torque control [32], aimed to give a detailed contribution for a detailed comparison between the two control techniques, emphasizing advantages and disadvantages. The performance of the two schemes is evaluated in terms of torque and current ripple, and transient response to step variations of the torque command. A new DTC scheme has been also presented in order to improve the performance of the basic DTC scheme. The conclusion is that DTC might be preferred for high dynamic applications, but, on the other hand shows higher current and torque ripple.

In 2002, Bhim Singh and Sumit Ghatak Choudhuri provided fuzzy logic based speed controllers for vector controlled induction motor drive [39] which presents a comparative study of the proportional integral, fuzzy logic; fuzzy pre compensated proportional integral, fuzzy proportional integral and hybrid speed controllers for vector controlled induction motor drives. It also presented the effectiveness of the various mentioned controllers for different modes of operation such as starting, speed reversal, load application and load removal.

In 2002, Thomas M. Wolbank et al, investigated the performance of both FOC and DTC with special emphasis to the operation around zero speed [33]. In that operating range it is assumed that without shaft sensor the flux position can only be determined by exploiting parasitic effects of the machine such as saturation or slotting. Sensor less schemes to estimate the spatial flux position control suffer however from a reduced accuracy in specific load ranges which degrades the performance of the drive when compared with shaft sensor.

In 2003, J. Luukko, M. Niemela and J. Pyrhonen, Estimation of the flux linkage in a direct torque controlled drive – an improved integration method is presented for the estimation of stator flux linkage for speed and position sensor less direct torque controlled ac machine drives [41]. The approach detects the incorrectness of the stator flux linkage estimate by observing the scalar product of the estimated stator flux linkage and the measured stator current. An error term is formed by extracting the ac part of the scalar product by a proper filtering. The improvements were shown to have a good dynamic performance, which widens the application range of the DTC- controlled drives.

In 2003, M. Cirrincione, M. Pucci and G. Vitale, presented a paper [40] which gives a technique when direct torque control algorithm is applied to induction motor drives supplied by 3-level diode clamped inverters. Results obtained with numerical simulations show that the employment of a 3-level inverter in a such a control scheme permits to obtain the same dynamical performances as those obtained with a 2-level inverter resulting far low torque and flux ripples as well as lower harmonic content in the stator current.

In 2006, R. A. Gupta, Rajesh Kumar and Borra Suresh Kumar provided a technique direct torque controlled induction motor drive with reduced torque ripple [43]. DTC allows direct independent control of flux linkage and electromagnetic torque by the selection of optimum inverter switching modes. But the possibility of generation of only six non-zero voltage vectors results in the production of ripples in the electromagnetic torque and stator flux linkages. In this paper a new torque ripple reduction strategy applicable for direct torque control is presented. This methodology involves the production of zero vectors along with a non-zero vector during a switching interval determined by the torque and flux errors as a result 12 combinations of non-zero voltage vectors and 2 zero voltage vectors are possible.

In 2006, Bhim Singh, Pradeep Jain, A. P. Mittal and J. R. P. Gupta proposed a practical approach to electric vehicle employing direct torque control [44]. The DTC based induction motor control for electric vehicle propulsion system discussed in this paper provides quick response, simple conFigureuration. the proposed scheme is capable of providing four quadrant operation along with regenerative braking with partial recovery of kinetic energy to charge the battery and thereby improving the overall efficiency of the system .

In 2007, N. Farid , B. Sebti , K. Mebarka and B. Tayeb presented a paper [45]on the analysis of field oriented control and direct torque control for sensor less induction motor drives. The main aim was to give a fair comparison between IFOC and DTC techniques. The synthesis of the study reveals a slight advantage of DTC scheme compared to IFOC scheme regarding the dynamic flux control performance. The DTC might be preferred for high dynamic applications, but, shows higher current and torque ripple. It quotes the choice of one or the other scheme will depend mainly on specific requirements of the applications.

In 2007, Sadegh Vaez-Zadeh and Ehsan Jalali developed an Induction motor drive system employing features of vector control and direct torque control [46]. In this paper a new control system is proposed for three phase high performance induction motor drives based on common basis provided for vector controlled and direct torque controlled induction motors. The control system enjoys the partially the advantages of vector control and direct torque control and avoids some of the implementation difficulties of either of the two control methods. An extensive comparative performance evaluation of a motor under the proposed control system confirms its effectiveness and partial superiority over either vector control or direct torque control, despite its relative structural simplicity.

AUTHOR'S CONTRIBUTION:

In this work, Vector Control and Direct Torque Control techniques have been simulated separately for various modes of operations namely, starting, speed reversal and load perturbation (load application and load removal). A comparative study between field oriented control (FOC) and direct torque control (DTC) is presented with the results obtained by simulation in MATLAB environment using simulink and power system blockset (PSB) toolboxes. The simulation and evaluation of both control strategies are performed on induction motors of two different ratings namely, 1HP and 30HP.

1.4 CONCLUSION

From the above discussion it can be concluded that the control of induction motor is very necessary as it is the common motor used in industrial motion control systems, as well as in main powered home appliances. Hence a well established induction motor drive which is simple, rugged, low cost and low maintenance can serve the required purpose. Many authors have published several research papers on the control techniques of induction motor. At the same time it is also necessary to control the output voltage of the inverter used in the drive. These control techniques are also explained in the next chapters of this thesis. The modeling and MATLAB implementation of the Vector Controlled Induction Motor Drive is also explained in detail in the third chapter and the modeling and MATLAB implementation of the Direct Torque Controlled is explained in the fourth chapter. Fifth chapter gives the comparison results obtained by simulation of both the techniques in MATLAB environment. Finally sixth chapter gives the conclusions and future scope of the work.

CHAPTER II

VECTOR CONTROLLED INDUCTION MOTOR DRIVE

2.0 INTRODUCTION

Vector Control technique has considerably widened the scope of application of cage induction motors. Now, this motor can be used even in those applications which were so far considered to be the exclusive domain of dc motors [3]. The separately excited dc motor has a linear control plant with a decoupled control structure which facilitates the independent control of flux and torque. Unlike this, the induction motor has a non-linear and highly interacting multivariable control plant. The method of vector control helps in transferring its dynamic structure in such a way that it behaves like a separately excited dc motor.

An induction motor can be used in the place of a dc motor as the induction motor is smaller, mechanically more rigid, less expensive, maintenance free as it does not have commutator and brushes. In the vector control method, an induction motor is controlled like a separately excited dc motor. It is well known that in the case of fully compensated dc motor, the stationary poles produced either with dc-excited field winding or permanent magnets establish a magnetic field in which the armature rotates. The current is supplied to the armature through brushes and commutator so the armature mmf axis is established at quadrature with the main field axis.

The orthogonal relationship between the field flux and armature mmf axis is independent of the speed of rotation and therefore, the electromagnetic torque developed by the fully compensated dc motor is proportional to the product of the field flux and armature current. With a constant field flux, the torque becomes directly proportional to the armature current and hence it can be altered as rapidly as the variation of armature current [4].

2.1 DC MOTOR ANALOGY

The analogy between the vector controlled induction motor and dc motor can be explained with the help of Figure 2.1. In a fully compensated dc motor, the torque is given by

$$T_e = K_{fdc} \times I_a \times I_f$$

Where, K_{fdc} is the torque constant of the dc motor, I_a is the armature current and I_f is the field current.

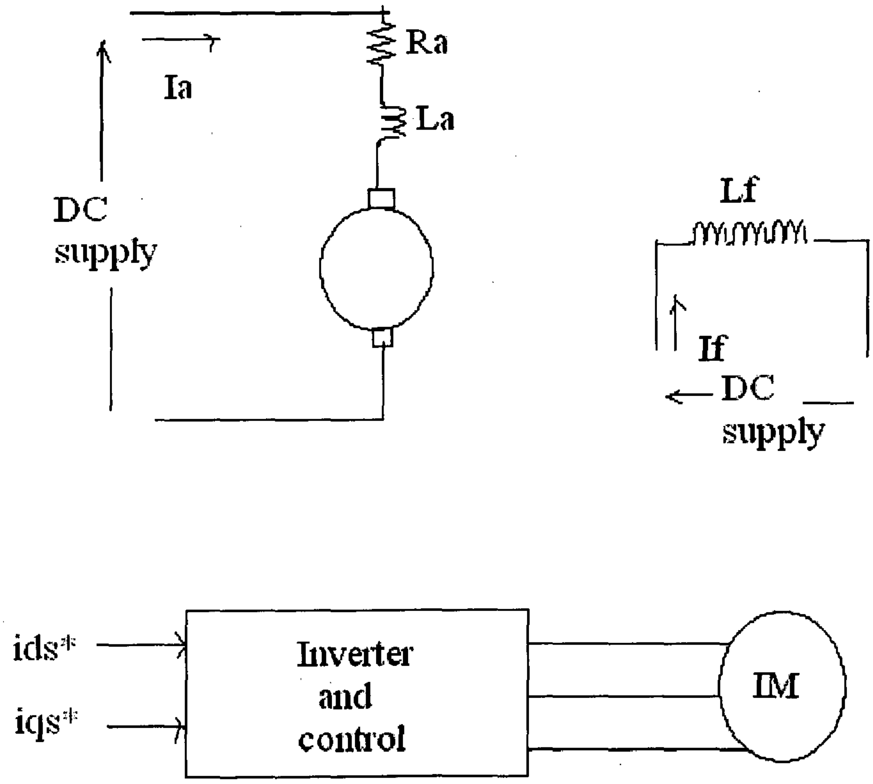

Figure 2.1 Induction motor analogy with DC motor in vector control

The angle (90 degrees) between two fields created by I_a and I_f is independent of both the speed of the motor and the load on its shaft. These two fields are considered as orthogonal or decoupled vectors and are termed as independent control variables. The current I_f and the corresponding field flux is decoupled from the armature current I_a, thus accomplishing the fast dynamic response through quick control of the developed electromagnetic torque in transient as well as steady state operating conditions of the dc motor. This mode of control can be made applicable to an induction motor by controlling it in a synchronously rotating reference frame where the sinusoidal variables of the motor become dc quantities [7].

The induction motor with two control inputs, I_{ds}^* and I_{qs}^* are the direct-axis component and quadrature-axis component, respectively, of the stator current and are expressed in synchronously rotating reference frame. In the vector control, I_{ds}^* is analogous to the field current I_f and I_{qs}^* is analogous to the armature current I_a of a dc motor. Therefore, the torque in an induction motor is expressed as

$$T_e = K_{fim} \times I_{qs}^* \times I_{ds}^*$$

Where K_{fim} is the torque constant of the induction motor.

The torque of the induction motor expressed in equation is identical to that of the dc motor. The angle (i.e., relation of orthogonality) between I_{ds}^* and I_{qs}^* is independent of both the speed of the motor and the load on its shaft, thereby fulfilling the condition of the vector control. For normal operating conditions below the base speed the current I_{ds}^* is kept constant and the torque is varied by changing the variable I_{qs}^*. Therefore, with vector control strategy, the induction motor behaves like a fully compensated separately excited dc motor. The currents I_{ds}^* and I_{qs}^* are controlled independently in the flux control loop and torque control loop respectively.

2.2 CONTROLLED QUANTITIES IN VECTOR CONTROL

Like in a dc motor, torque is produced in the induction motor by the electromagnetic interaction of current in the rotor conductor and the magnetic flux density in the air gap to which the rotor is subjected. But in three phase induction motor of squirrel cage type, all the electrical inputs are given to stator only [5]. The currents flowing on the stator coils are primarily responsible for: inducing currents in the rotor and creating magnetic field in the airgap.

The technique of vector control is based on a method of separating out these basic two functions of the stator current. In a vector control induction motor, the current flowing in the stator phase is the sum of two components at every instant – one component known as field component responsible for producing the magnetic field in the airgap and the other component known as torque component responsible for inducing the torque by inducing the corresponding current in the rotor conductors.

The necessary computation for determining the torque component and the field component of each stator phase current is done online by suitably programmed computing blocks (MATLAB – Simulink blocks are used in this work). Then a closed loop controller compares the torque component against the demanded torque. The demanded torque is the reference input provided by an outer speed control loop. This loop determines the necessary change to make he torque component equal to the demanded torque. In a similar way there is a control loop for field control. This loop compares the computed value of the field component against the field reference. The field reference has a constant value at all speeds below the base speed as in the case of a DC motor. To implement field weakening above the base speed, the reference input to the field controller is reduced. In either case, the field control loop compares the computed value of the actual field component against the field reference. This loop determines the change, if any, necessary to make the actual field component equal to the commanded value.

In this manner, the torque control loop determines the required change in the torque component and the field control loop determines the change required in the field component. These two individual values of the required changes are added to get the total change required in the output current of the inverter. The inverter switching times are appropriately modified to implement the required total change.

2.3 CLASSIFICATION OF VECTOR CONTROL SCHEMES

In a broader sense there are two criteria which can be used for classifying the vector control of an induction motor. The first depends upon the method used in determining the rotor flux vector. The second one depends upon the selection of reference frame for facilitating the control of the motor. There are two methods of determining the rotor flux vector which are known as direct and indirect methods of vector control. Similarly, the vector control can also be classified depending upon fixation of reference frame. Thus one can have stator flux oriented control, rotor flux oriented control or the magnetizing flux oriented control [3]. Generally in induction motors, the rotor flux oriented control is preferred. This is due to the fact that by aligning the synchronously rotating reference frame with the rotor flux, the vector control structure becomes simpler and the dynamic response of the drive is observed to be better than any other alignment of the synchronously rotating reference frame.[2]. Vector control technique can be classified into (a) direct vector control and (b) indirect vector control [3,7]. These methods are described in the following sections.

2.3.1 DIRECT VECTOR CONTROL METHOD

Direct field-oriented control, determines the magnitude and position of the rotor flux vector by direct flux measurement or by a computation based on terminal conditions [4,7]. Direct vector control also called as flux feedback control, is a method in which the required information regarding the rotor flux (magnitude and alignment) is obtained by means of direct flux measurement or estimation. Direct flux measurement is achieved by using Hall effect sensors, search coils, tapped stator windings of the machine etc. Alternatively, rotor flux can also be achieved by use of flux model, which is based on the motor parameters and the electrical input variables. The major disadvantage of the method is the need for a number of sensors. This prevents the scheme from being used in motors, which are already in use and add to the manufacturing cost of new motors. Moreover, fixing up of flux sensors become a tedious job. Problems such as drift because of temperature, poor flux sensing at lower speeds also persist. In the direct vector control scheme the flux sensors are installed for acquisition of rotor flux vector. The quantities generated from flux sensors are used in the outer loop of the drive control structure. Alternatively, in place of the flux sensor, the flux model can also be used for which the stator currents and voltages become the feedback signals and the rotor flux angle is given as its estimated output.

2.3.2 INDIRECT VECTOR CONTROL METHOD

Existence of related problems in case of direct vector control leads to the use of indirect vector control. In indirect vector control technique, the rotor flux vector position is computed from the speed feedback signal of the motor [4,7]. The indirect method vector control eliminates the need of using flux sensors or flux model. However, it requires an accurate measurement of the shaft position in order to determine the precise position of the rotor flux vector. The difference between the reference speed and the rotor speed is fed to the controller. The controller computes the slip frequency that on addition to the feedback motor speed provides the speed of the rotor flux vector from which the flux can be computed. Both of these (w_r and w_2^*) are summed up and integrated to get the flux angle ψ and then this angle ψ is used for the two phase rotating i_{ds}^* and i_{qs}^* to three phase stationary i_{as}^* , i_{bs}^* and i_{cs}^* . These reference currents are reproduced in the motor winding by the current controlled inverter. Indirect vector control technique eliminates most of the problems, which are associated with the flux sensors as the controller is now free from rotor flux sensing. Between these two methods of vector control the indirect vector control, has been found to be simple and easy to implement.

Therefore, the method adopted in this work is indirect vector control or indirect rotor flux oriented control. The in phase component of the stator current space vector in the synchronously rotating reference frame (SRRF) is aligned with the excitation current for rotor flux (in other words aligned with the rotor mmf vector). This component of the stator current vector is responsible for production of flux. Similarly, the quadrature component of the stator current space vector is responsible for the production of torque. Henceforth, such a control technique provides a substitute of a separately excited dc motor using a three-phase squirrel cage induction motor in variable speed applications.

USE OF REFERENCE FRAME

For the purpose of control, the direct and quadrature –axis stator currents are obtained in the corresponding reference frame. These are similar to the field and armature currents of the separately excited dc motor. It is known that when an induction motor is supplied by impressed stator voltages, the stator flux oriented control scheme is simpler than the rotor flux oriented control scheme. While in an induction motor supplied by impressed stator currents, the implementation of the rotor flux oriented control is simpler than either the stator flux oriented control or the magnetizing flux oriented control. It is possible to implement stator flux and magnetizing flux oriented control in case of wound rotor induction motor. In case of cage motor the rotor flux oriented control is usually employed due to the fact that in rotor reference frame the vector controlled drive system is made to have a relatively simpler structure and its dynamic performance and linearity in the torque slip plane is superior to any other reference frame of vector control. Therefore, in this work rotor flux reference frame is used.

2.4 SPEED CONTROLLER

In this work, the proportional integral speed controller has been used; speed controller is followed by a limiter. Generally the speed error, which is the difference of reference speed and actual speed, is given as input to the controller. The speed controller processes the speed error and gives torque value as an output [39]. Then the toque value is fed to the limiter which gives the final limited reference torque. The speed error at any *n*th instant of time is given as:

$$\omega_{r\,e(n)} = \omega_{r(n)}^{*} - \omega_{r(n)}$$

where $\omega_{r(n)}^{*}$: reference speed of the motor

$\omega_{r(n)}$: actual speed of the motor

2.4.1 Proportional Integral (PI) speed controller

The PI speed controller is the simplest speed controller compare to any other speed controller. The general block diagram of the PI speed controller is shown in Figure 2.2. The output of the speed controller at nth instant is expressed as follows:

$$T_{(n)} = T_{(n-1)} + K_p\{\omega_{re(n)} - \omega_{re(n-1)}\} + K_i \omega_{re(n)}$$

where $T_{(n)}$: torque output of the speed controller at the n^{th} instant.

K_p and K_i are proportional and integral gain constants

$\omega_{re(n-1)}$: speed error at (n-1)th instant

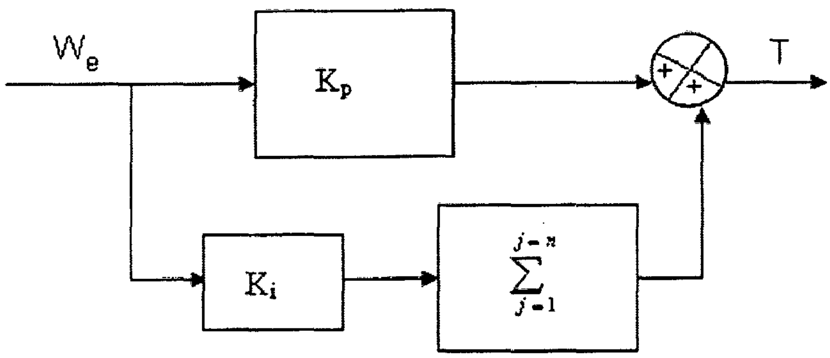

Figure 2.2 Block diagram of PI speed controller

The PI speed controller gain parameters shown in the block diagram and the equation are selected by trial and error basis by observing their effects on the response of the drive. The numerical values of these controller gains depend on the ratings of the motor and they are presented in the Appendix-B for the motor drive system used in this work.

The MATLAB/simulink model diagram of the PI controller in discrete time is shown in Figure 2.3 as seen in Figure, using the proportional and integral gain parameters namely, k_p and k_i

respectively along with the limiter, the reference torque is calculated in accordance to the motor rating used.

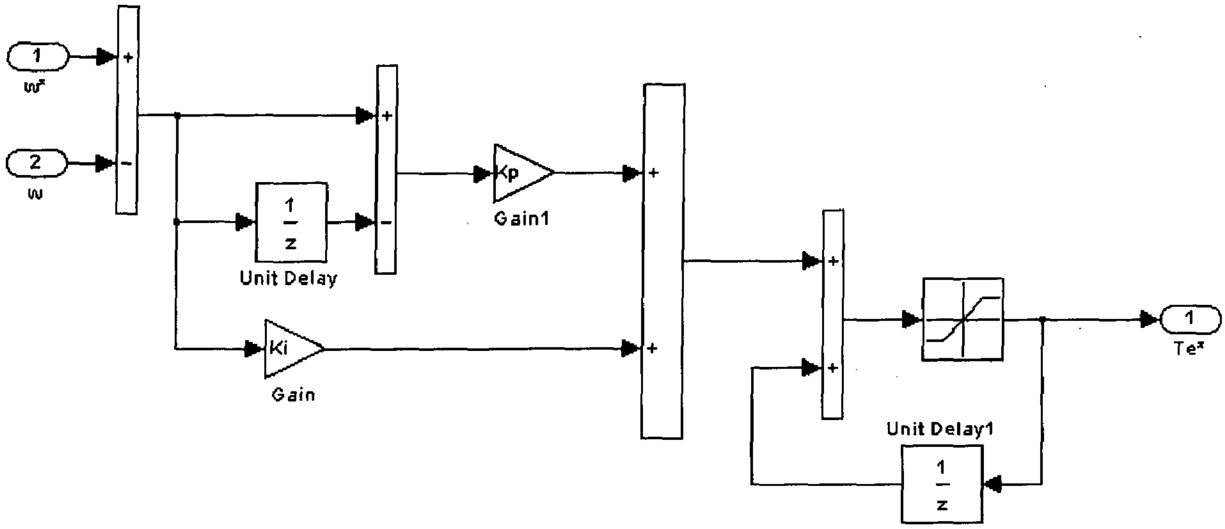

Figure 2.3 MATLAB/simulink model diagram

2.5 CONCLUSION

In this chapter, the basics of the vector control technique have been discussed and the classification of vector control technique is explained. The induction motor is compared with the dc motor. So that it is possible to control the induction motor similar to the separately excited dc motor i.e., by decoupling the flux producing component and the torque producing component. The MATLAB simulink model has been explained for the PI speed controller.

CHAPTER III

MODELLING AND SIMULATION OF VECTOR CONTROLLED INDUCTION MOTOR DRIVE

3.0 INTRODUCTION

The mathematical model used in this simulation work consists of equations expressed in terms of real operating variables of the motor such as winding currents and voltages. Therefore, the simulated results obtained from this model represent the response of the real time vector control induction motor drive. In order to facilitate an easy understanding of the model and the working of VCIMD, a detailed description of components of drive system is presented. As the vector controlled cage motor drives are normally fed from a current controlled voltage source inverter, the working of this particular source is also described.

3.1 VECTOR CONTROL SYSTEM

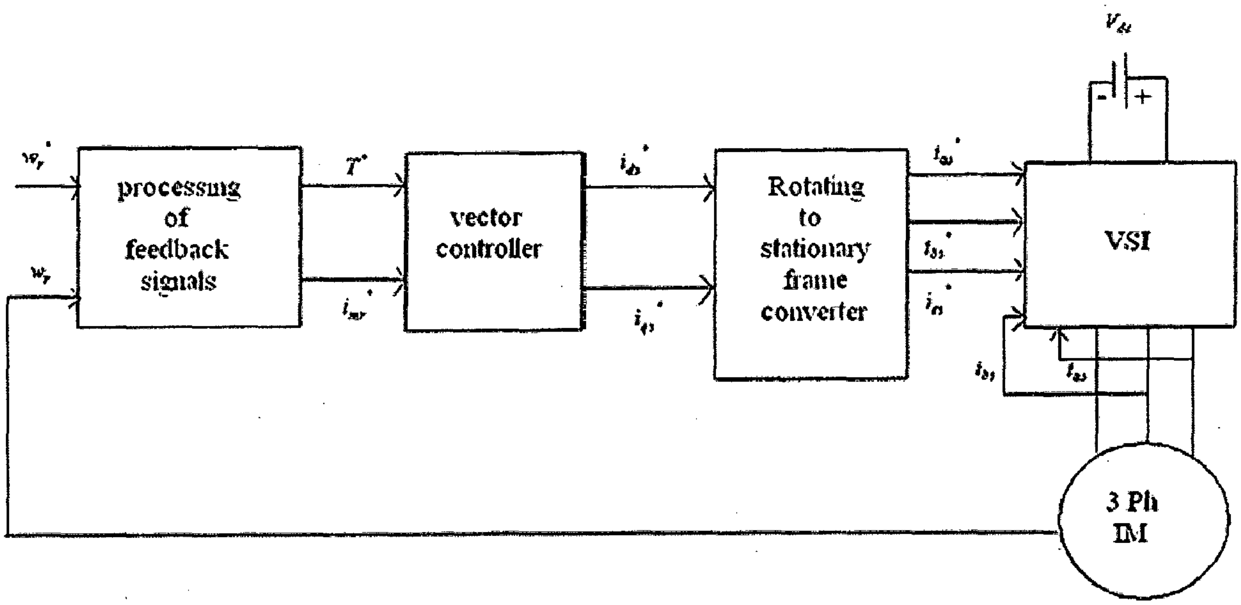

Figure 3.1 Schematic Diagram for Vector Controlled Induction Motor

Figure 3.1 shows the schematic of the proposed drive system which uses current controlled voltage source inverter to supply the 3 phase squirrel cage induction motor. The working of the system can be explained by identifying the flow of signals based on the input-output relationship of individual components. For the purpose of control, two outer loops, namely the torque control loop and flux control loop and an inner loop which controls the current are easily recognized. The flow of different signals through these loops is described with a view to explaining the operation (starting, speed reversal, load perturbation and steady state running at the set speed) of the drive. It is observed from Figure, that the rotor speed w_r is sensed with the help of a speed sensor and the signal so obtained is compared with the signal of reference speed w_r^*. This results in an error signal w_e which is processed in the speed controller of the torque control loop. The output of the speed controller is fed to the limiter which gives its output T^*, which is considered as the reference torque of the drive.

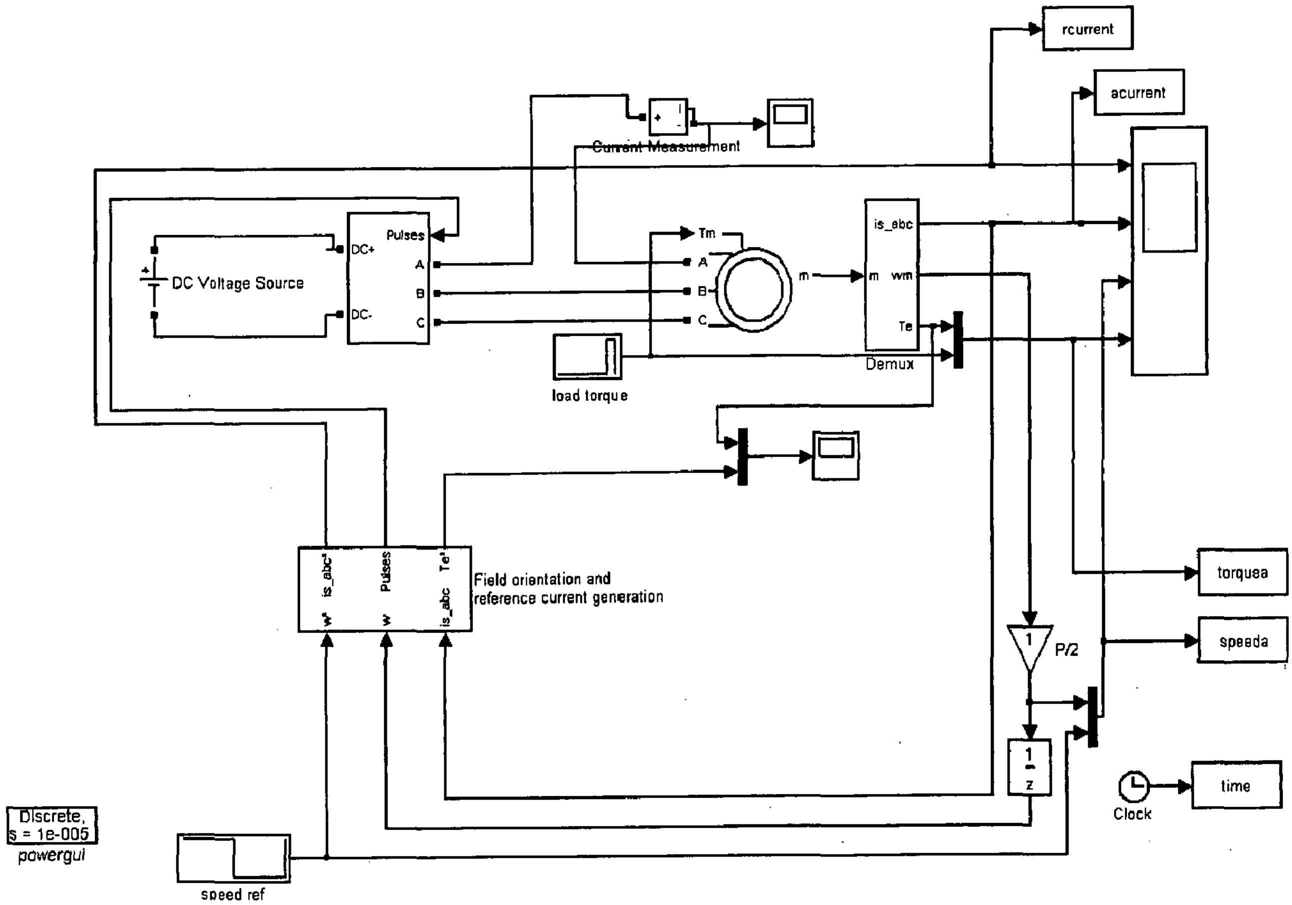

Figure 3.2 MATLAB/simulink model of vector controlled induction motor drive

3.2 VECTOR CONTROL TECHNIQUE

The field-weakening controller receives the speed signal (w_r) as an input signal and provides reference value of the excitation current (i_{mr}^*) as an output signal. Therefore, the two signals T^* and the i_{mr}^* are the reference signals for the vector controller [3]. In the vector controller the d-axis component (i_{ds}^*) and the q-axis component (i_{qs}^*) of the stator current signals are computed which are responsible for the flux control and the torque control respectively. The slip frequency signal (w_2^*) is also computed in vector controller to evaluate the flux angle. The flux angle (ψ) is computed using slip frequency (w_2^*), rotor speed (w_r) and sampling period (ΔT). These signals of flux (i_{ds}^*) and torque (i_{qs}^*) control are in the synchronously rotating reference frame and these are transformed into stationary reference frame three-phase currents (i_{as}^*, i_{bs}^* and i_{cs}^*) using flux angle (ψ).

For current controlled VSI fed vector-controlled induction motor, the reference currents (i_{as}^*, i_{bs}^* and i_{cs}^*) and sensed winding currents (i_{as}, i_{bs} and i_{cs}) are fed into the pulse width modulated (PWM) current controller. The current error signals are amplified and used as the modulating signals for the PWM controller. A triangular carrier waveform is generated at the required switching frequency (f_s). The point of intersection of the triangular carrier wave and the modulating signals acts as the point of state changeover for the resulting PWM signals, which are fed to the driver circuit of the IGBT of VSI feeding an induction motor.

Figure 3.3 shows the structure of VCIMD. It consists of the speed controller, the field weakening controller, the vector controller, two phase rotating frame to three phase stationary frame converter, PWM current controller, CC-VSI and three phase squirrel cage induction motor. The functions of these blocks are described below:

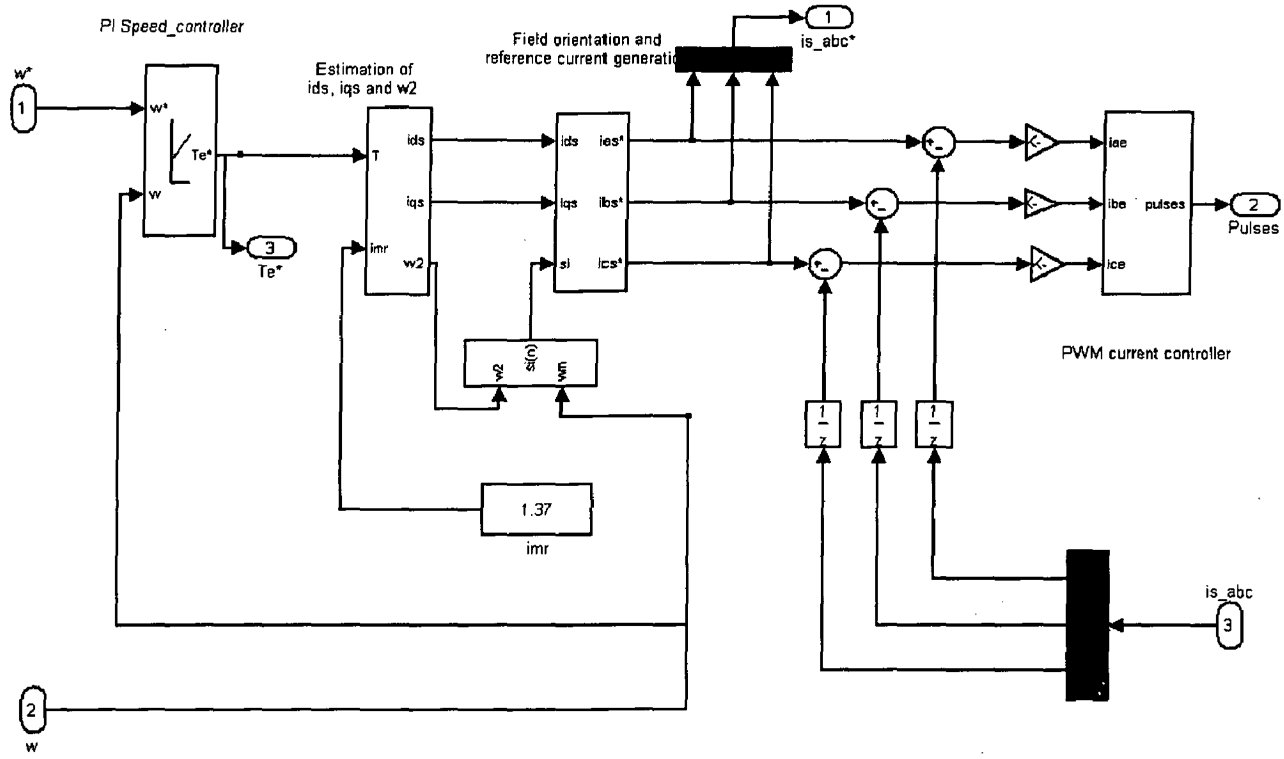

Figure 3.3 MATLAB model of vector control technique

3.2.1 SPEED SENSOR

It measures the motor speed, since in the indirect vector control the accurate measurement of position of rotor flux vector is required which may be measured with the help of a high resolution and high precision speed sensor. Normally, for this purpose, shaft encoders are used for the closed loop vector control of the cage induction motor drive.

3.2.2 SPEED CONTROLLER

The measured speed (w_r) is compared with the set reference speed (w_r^*) in the error detector and the resulting output is known as speed error (w_e) is processed in the speed controller. The output of the speed controller is the control signal for torque command (T). The command input (w_e) may be positive or negative depending upon the values of set reference speed and the motor shaft speed. Speed error (w_e) is processed in the speed controller which may be of different types depending upon the required dynamic performance of the drive. In the present work proportional integral (PI) controller is used.

3.2.3 LIMITER

When the drive operates in the transient conditions such as starting, reversing or load application or load removal the speed controller output (T) may be very high value to achieve the steady state condition of the drive as fast as possible, as a result, the controller output signal (T) may become quite high and in some cases it may become higher than the breakdown torque of the motor. Such a situation may be rather dangerous for the motor and may take the drive into instability. In order to avoid such circumstances, it becomes very much necessary to apply certain limit on the output of the speed controller. The output of the speed controller after the limit is considered as the reference torque (T*) to the vector controller and used to obtain the value of stator current torque component of the stator current space vector. As a result, the limit on the torque also ensures overcurrent protection to the drive. Whenever reference speed change or there is an application of load torque on the motor shaft, the speed controller output is limited to a maximum permissible value (T*). Therefore, the limiter on the speed controller output provides an inherent stability to the closed loop speed control system.

3.2.4 FIELD WEAKENING CONTROLLER

The field weakening operation of a VCIMD is similar to the field control of a separately excited dc motor. This operation is implemented when the drive speed is controlled above the base speed. The input to the field-weakening controller is the feedback speed of the motor. The output of the controller is the excitation current. Below the base speed the excitation current remains constant. Above the base speed the excitation current varies in inverse proportion to the speed.

$i_{mr}^*(n) = i_m$ if $\omega_r(n)$ < base speed of motor

$i_{mr}^*(n) = K_f\, i_m / \omega_r(n)$ if $\omega_r(n)$ >= base speed of motor

where K_f : flux constant

i_{mr}^* : excitation current, i_m : magnetizing current, $\omega_r(n)$: feedback speed of the motor drive, ω_b : base speed of the motor drive.

3.3 MATHEMATICAL ANALYSIS OF VECTOR CONTROLLER

The output of the speed controller after limiting is taken as the reference torque (T*) and output of field weakening controller (i_{mr}^*) is taken as reference flux for the vector controller. These two command signals are taken as input to the vector controller for calculating the torque component (i_{qs}^*) and the flux component (i_{ds}^*) of the stator current (i_s^*) and also to calculate the slip frequency (w_2^*). The torque and flux current components (i_{ds}^* and i_{qs}^*) are the respective decoupled components of the stator current (i_s^*) in the synchronously rotating reference frame.

Estimation of i_{ds}^*, i_{qs}^* and ω_2^*

The modeling of vector controller block is the heart of the entire modeling of the vector controlled induction motor drive. This section calculates the direct and the quadrature axis stator current components (i_{ds}^* and i_{qs}^*) in the synchronously rotating reference frame (SRRF) aligned with rotor inclined at flux angle (ψ) with respect to stationary reference frame (SRF).

Mathematically, the equations for calculating these two components of the current in the discretized form are stated as follows [2,4,39]:

$$i_{ds}^*(n) = i_{mr}^*(n) + \tau_r \frac{di_{mr}^*}{dt}$$

$$i_{qs}^*(n) = \frac{T^*(n)}{Ki_{mr}^*(n)}$$

$$w_2^*(n) = \frac{i_{qs}^*(n)}{\tau_r i_{mr}^*(n)}$$

where τ_r is the rotor time constant defined as

$$\tau_r = \frac{L_r}{R_r}$$

$$K = \left(\frac{2}{3}\right)\left(\frac{P}{2}\right)\left(\frac{M}{1+\sigma_r}\right)$$

P is the number of poles, i_{ds}^* (n) and i_{qs}^* (n) refer to flux and torque components of stator current at nth instant, w_2^* (n) refer to nth instant reference slip frequency, M is the mutual inductance, σ_r is the rotor leakage factor and L_r is the rotor self inductance and is defined as below $L_r = L_{lr} + L_m$

$$L_r = (1+\sigma_r)M \qquad \sigma_r = \left(\frac{Lr}{M}\right) - 1$$

where $M = \frac{3}{2}L_m$, R_r : rotor resistance , L_m is the magnetizing inductance.

The MATLAB model for the estimation of i_{ds}^*, i_{qs}^* and ω_2^* is shown in Figure. 3.4

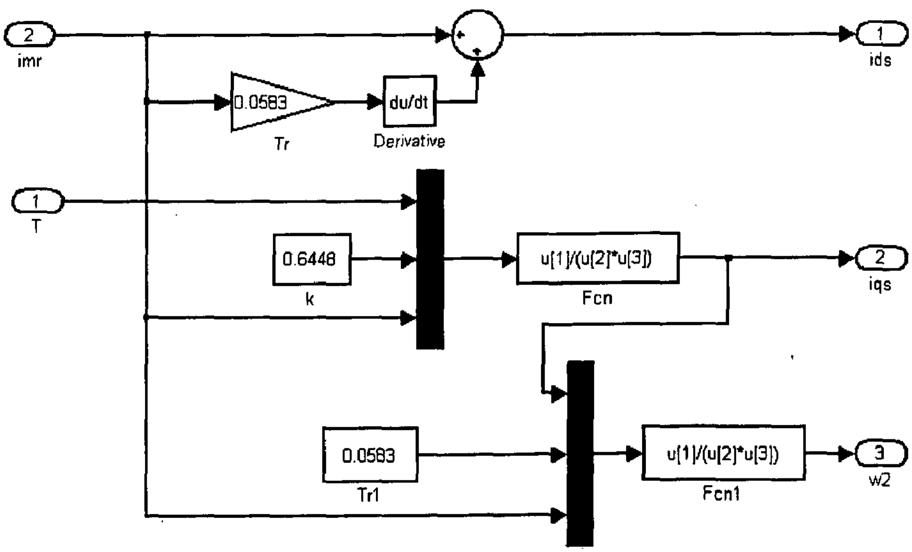

Figure 3.4 Estimation of i_{ds}^*, i_{qs}^* and ω_2^*

3.4 TWO PHASE ROTATING FRAME TO THREE PHASE STATIONARY FRAME CONVERTER

Two phase rotating frame to three phase stationary frame converter transforms the two decoupled components of stator current namely , i_{ds}^* and i_{qs}^* in synchronously rotating frame into three phase currents namely i_{as}^*, i_{bs}^* and i_{cs}^* in three phase stationary reference frame. The conversion process requires the flux angle (ψ), which is calculated by integration of synchronous speed. Synchronous speed is obtained by addition of slip speed (w_2^*) and motor speed (w_r).

Two-Phase rotating to three phase stationary reference frame converter can be modeled as follows:

$$i_{as}^* = -i_{qs}^* \text{ Sin } \psi + i_{ds}^* \text{ Cos } \psi$$

$$i_{bs}^* = [(-\text{Cos } \psi + \sqrt{3} \text{ Sin } \psi)\, i_{ds}^* (\frac{1}{2})] + [(\text{Sin } \psi + \sqrt{3} \text{ Cos } \psi)\, i_{qs}^* (\frac{1}{2})]$$

$$i_{cs}^* = -(i_{as}^* + i_{bs}^*)$$

Where i_{ds}^* and i_{qs}^* refer to decoupled components of the stator circuit i_s^* in two-phase system w.r.to rotor reference frame and i_{as}^* , i_{bs}^* and i_{cs}^* are three phase currents in stator reference frame.

3.5 SINUSOIDAL PULSE WIDTH MODULATED (PWM) CURRENT CONTROLLER

Current control methods play an important role in power electronic circuits, particularly in current regulated PWM inverters which are widely applied in ac motor drives and continuous ac power supplies where the objective is to produce a sinusoidal ac output. The main task of the control system in current regulated inverters is to force the current vector in the three phase load according to a reference trajectory. In order to operate the three phase induction motor into vector controlled mode, sensed three phase stator currents (i_{as}, i_{bs} and i_{cs}) are to be controlled close to the three phase reference currents(i_{as}^*, i_{bs}^* and i_{cs}^*). Such a control technique to control the two sets of these phase currents is called current control. Sinusoidal PWM current control is one of the current control technique wherein the switching frequency can be kept constant.

$$i_{aes}^*(n) = i_{as}^*(n) - i_{as(n)}$$

$$i_{bes}^*(n) = i_{bs}^*(n) - i_{bs(n)}$$

$$i_{ces}^*(n) = i_{cs}^*(n) - i_{cs(n)}$$

The error between the reference three phase currents and the sensed three phase currents are calculated and the set of current error are amplified and the resulting output is taken as modulating signals which are compared with a carrier triangular wave of required switching frequency(f_s). For a given phase, the instant at which the modulating signal and triangular carrier wave intersect each other is the instant at which the switching signal for that respective phase changes its logical value between switch off and switch on. The switch off or switch on PWM logic generated for all the three phases are given to the gate driver of IGBT of VSI.

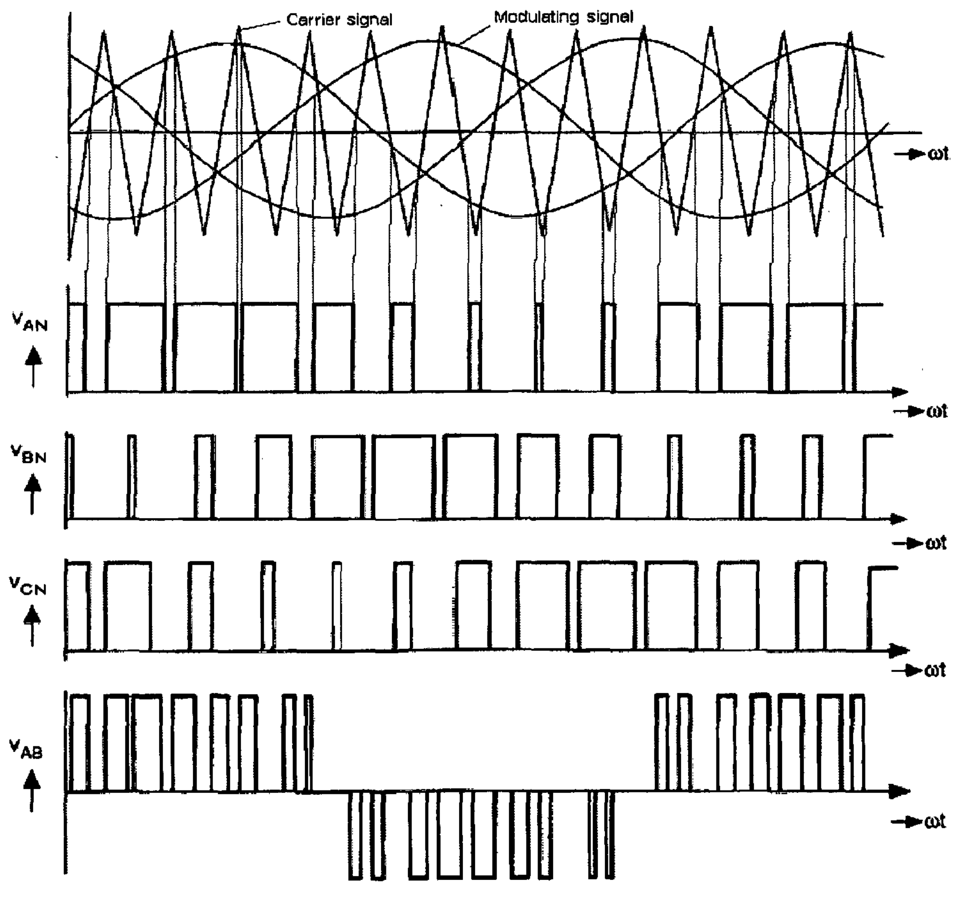

Figure 3.5 Three phase sinusoidal PWM [9]

3.6 HYSTERESIS CURRENT CONTROLLER:

Among the various PWM technique, the hysteresis band current control is used very often because of its simplicity of implementation. Also, besides fast response current loop, the method does not need any knowledge of load parameters. The basic implementation of hysteresis current control is based on deriving the switching signals from the comparison of the current error with a fixed tolerance band. This control is based on the comparison of the actual phase current with the tolerance band around the reference current associated with that phase.

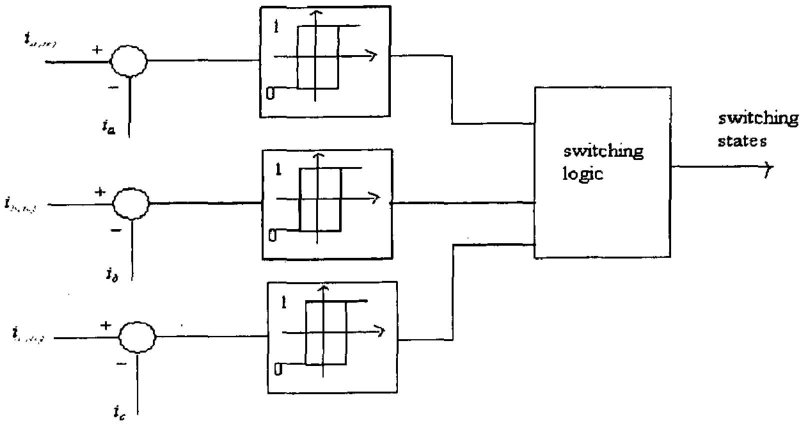

Figure 3.6 Hysteresis Controller Technique

VOLTAGE SOURCE INVERTER (VSI)

The three phase voltage source inverter (VSI) comprises of a bridge configuration of six IGBT switches with respective freewheeling diodes. Sinusoidal PWM current controller operates the VSI in current control mode [47]. The controlled VSI forms a variable voltage source at varying frequencies for the three phase induction motor. The output voltage of the inverter depends on the combination of switching functions SF_A, SF_B and SF_C. The three phase voltages can be expressed as:

$$V_{as}=V_{dc}(2SF_A - SF_B - SF_C)/3$$

$$V_{bs}=V_{dc}(-SF_A+2SF_B - SF_C)/3$$

$$V_{cs}=V_{dc}(- SF_A- SF_B+2SF_C)/3$$

3.7 THREE PHASE SQUIRREL CAGE INDUCTION MOTOR

The three phase squirrel cage induction motor is the main component of the drive. The motor runs at the set reference speed (w_r^*) in the required direction and converts supplied electrical energy into mechanical energy, robust in construction and free from maintenance.

Mathematical Model of Induction Machine:

The squirrel cage induction motor is modeled using d-q theory in the stationary reference frame so that we get less number of equations and the analysis becomes easy. The voltage-current relationship in the stationary reference frame of the induction motor in terms of the d-q variable is expressed as [2][47]:

$$[v] = [R][i] + [L]p[i] + \omega_r[G][i]$$

Therefore by simplifying, the current derivative vector can be expressed as follows:

$$p[i] = [L]^{-1}\{[v] - [R][i] + \omega_r[G][i]\}$$

where 'p' is the differential operator (d/dt) and 'ω_r' is the rotor speed in electrical 'rad/sec'.

Current and Voltage vectors are given as follows:

$$[i] = \begin{bmatrix} i_{qss} & i_{dss} & i_{qrs} & i_{drs} \end{bmatrix}^T$$

$$[v] = \begin{bmatrix} v_{qss} & v_{dss} & v_{qrs} & v_{drs} \end{bmatrix}^T$$

where v_{qss} and v_{dss} are the q- and d-axis voltages applied across the stator windings referred to the stator and v_{qrs} and v_{drs} are the q- and d-axis voltages across the rotor windings referred to the stator. As the rotor bars are short circuited in a squirrel cage induction motor, the voltages v_{qrs} and v_{drs} are zero. Similarly the currents are also defined. [L] is the inductance vector, [R] is the resistance vector and [G] is the rotational inductance matrix and have the usual meaning. All the voltages and currents are expressed in the stationary reference frame.

The torque balance equation is stated as:

$$p\omega_r = \left(\frac{P}{2}\right)\frac{(T_e - T_L)}{J}$$

Where T_L : load torque on motor including friction and windage losses,

T_e : developed electromagnetic torque by the motor.

3.8 RESULTS

The simulation has been performed using actual parameters of 1 HP and 30 HP induction motor fed by an IGBT inverter for starting, speed reversal, speed re-reversal and load perturbation and the results using Sinusoidal PWM and Hysteresis Current Controller have been shown in Figure 3.7 - 3.10 resp.

SINUSOIDAL PWM:

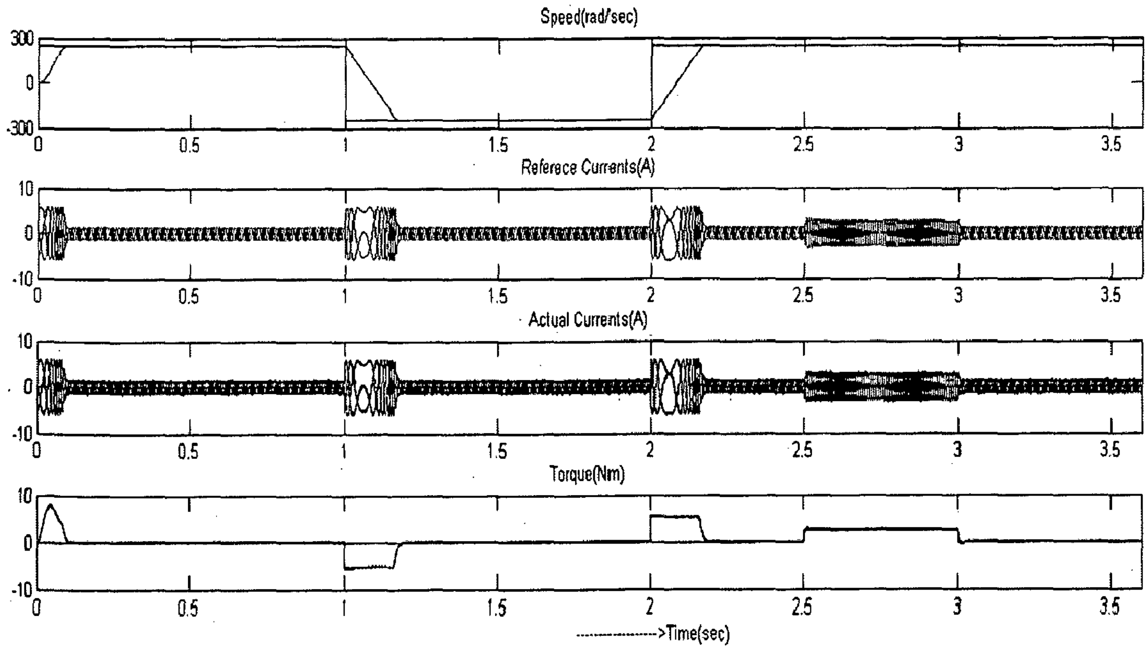

Figure 3.7 Starting, Speed Reversal, Speed Re-reversal and Load Perturbation Response of 1HP Vector Controlled Induction Motor Drive system.

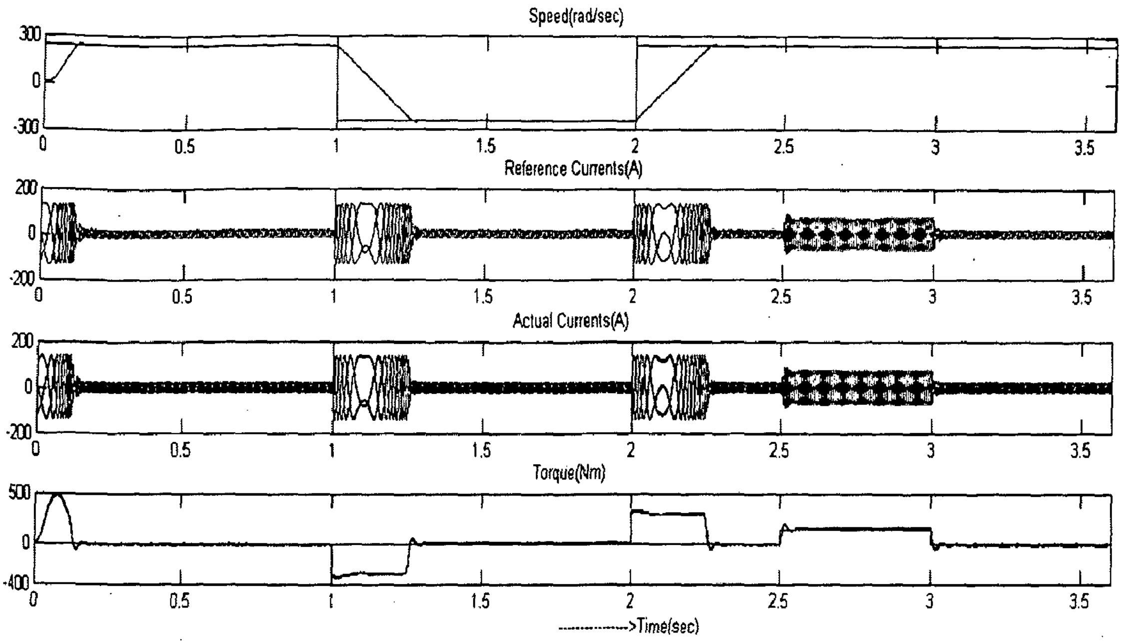

Figure 3.8 Starting, Speed Reversal, Speed Re-reversal and Load Perturbation Response of 30HP Vector Controlled Induction Motor Drive system.

HYSTERESIS CURRENT CONTROLLER:

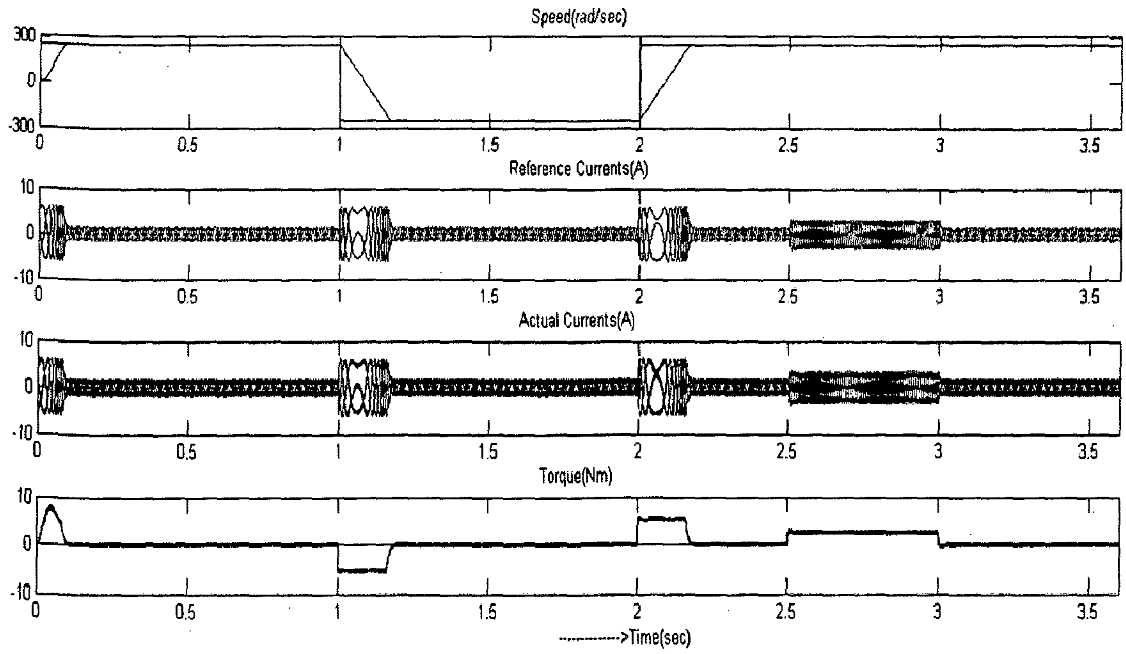

Figure 3.9 Starting, Speed Reversal, Speed Re-reversal and Load Perturbation Response of 1HP Vector Controlled Induction Motor Drive system Hysteresis Current Controller.

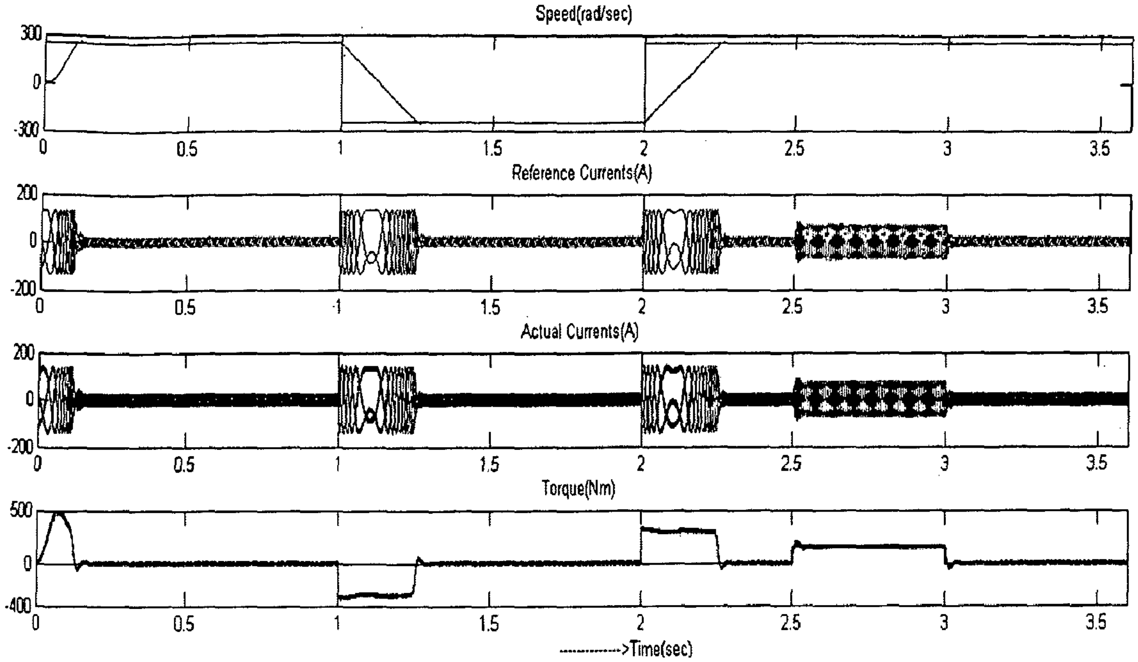

Figure 3.10 Starting, Speed Reversal, Speed Re-reversal and Load Perturbation Response of 30HP Vector Controlled Induction Motor Drive system using Hysteresis Current Controller.

3.9 CONCLUSION

In this chapter the main concept of the vector control technique of the induction motor drive with the mathematical analysis and the schematic block diagram has been presented. It is also discusses regarding the speed sensor, speed controller, the torque limiter and the field weakening method. This chapter also provides the technique of sinusoidal PWM and the hysteresis current controller technique and the results with both the techniques have been provided and the mathematical analysis of the 3 phase squirrel cage induction motor is also presented.

CHAPTER IV

MODELING AND SIMULATION OF DIRECT TORQUE CONTROLLED INDUCTION MOTOR DRIVE

4.0 INTRODUCTION

The control of a.c. drives is more difficult because, in ac machines, both the phase angle and the modulus of the current have to be controlled, or in other words the current vector has to be controlled. This technique is well known vector control technique [7]. In the same way in Direct Torque Control, it can be possible to control the induction machine by controlling the variables of the voltage vector that means by controlling the phase and modules of the voltage vector, the control of a.c. machine is achieved. But in d.c. machines, the orientation of the field flux and armature mmf is fixed by the commutator and the brushes, while in a.c. machines the field flux and the spatial angle of the armature mmf require external control. In the absence of this control, the spatial angles between the various fields in a.c. machines vary with the load and results into unwanted oscillating dynamic response.

4.1 DIRECT TORQUE AND FLUX CONTROL

The field-oriented control which guarantees high dynamic and static performance like dc motor drives, has become very popular and has constantly been developed and improved. When , in the mid 1980's, it appeared that control systems would be standardized on the basis of the FOC philosophy, there appeared the innovative studies of Depenbrock and Takahashi and Noguchi, which depart from the idea of co-ordinate transformation and the analogy with dc motor control. These innovators proposed to replace motor decoupling via nonlinear coordinate transformation with self control, which goes together very well with on-off operation of inverter power devices. In contrast to FOC systems where the flux and torque is controlled by acting on the direct and quadrature current vector components, respectively, Direct Torque and Stator Flux has been developed to regulate flux and torque directly, while currents and voltages are regulated indirectly.

In recent years the high Performance induction machine drives market has been dominated by the Field Oriented Control (FOC) methods. Complexity is a well known weakness of FOC technique. Many studies have been developed to find out different solutions for the induction motor torque control, reducing the complexity of FOC schemes and allowing precise and quick responses. Among these D.T.C. is very popular.

In principle classical DTC selects one of the six voltage vectors and two zero voltage vectors generated by a VSI in order to keep stator flux and torque within the limits of two hysteresis bands. The right application of this principle allows a decoupled control of flux and torque without the need of coordinate transformations, PWM pulse generation and current regulators. DTC-based drives selects the inverter switching states using the inverter-switching table, neither current controllers nor pulse-width modulation(PWM) modulator is required, thereby providing fast torque response[6].

The main features of the DTC technique is direct control of flux and torque, indirect control of stator currents and voltages, stator fluxes and voltages are approximately sinusoidal and dynamic performance.

The advantages of DTC are absence of coordinate transformations, absence of several controllers, only the sector where the flux linkage space vector is located has to be determined and not the actual flux-linkage space vector position and minimum time for the torque response.

The main disadvantages of the conventional DTC are the requirement of flux and torque estimators, the high ripple content in the torque and current.

4.2 Direct Torque Control System

A block diagram of a direct torque control system for an induction motor is shown in the Figure. 4.1. In this system, the control reference frame is stationary (fixed to stator). The motor torque and stator flux amplitudes are controlled by two independent hysteresis controllers. The feedback signals, T_e and ψ_r are computed from stator voltages and currents [7]. In a direct torque controlled (DTC) induction motor drive, supplied by a voltage source inverter, it is possible to control directly the stator flux linkage and the electromagnetic torque by the selection of optimum inverter switching modes. The selection is made to restrict the flux and torque errors within respective flux and torque hysteresis bands, to obtain fast torque response, low inverter switching frequency and low harmonic losses.

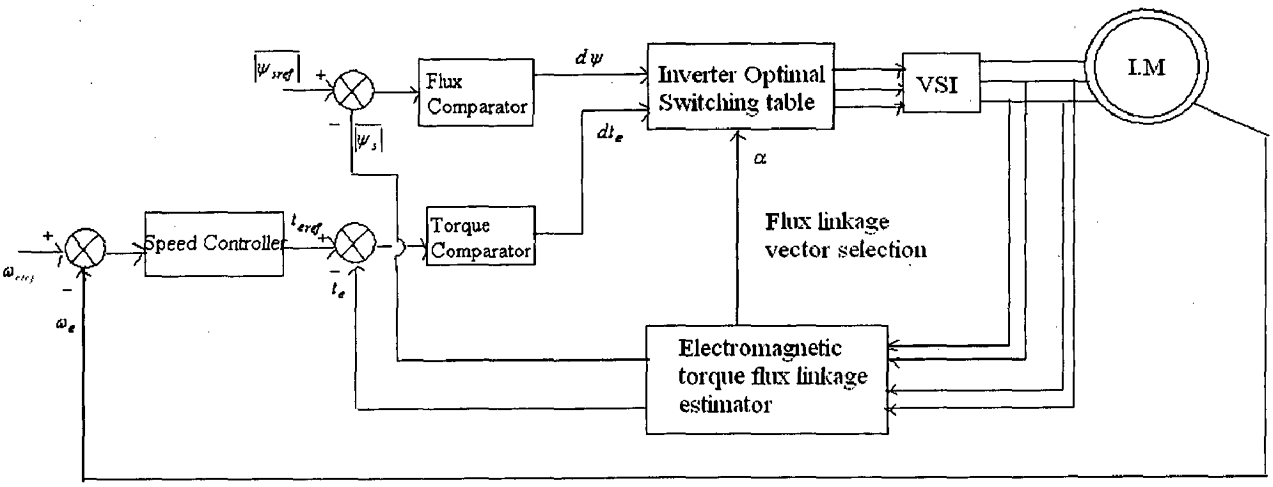

Figure 4.1 Schematic of stator flux based DTC induction motor drive with VSI [7]

The instantaneous electromagnetic torque of the three phase induction machines is proportional to the cross-vectorial product of the stator flux-linkage space vector and the stator-current space vector as follows

$$t_e = \frac{3}{2} P \bar{\psi}_s \times \bar{i}_s$$

Where $\overline{\psi_s}$ the stator flux-linkage space vector and $\bar{i_s}$ is the stator current space vector both space vectors are expressed in the stationary reference frame and P is the number of pairs of poles. By considering that $\overline{\psi_s} = |\psi_s| e^{j\rho_s}$, where ρ_s is the angle of the stator flux linkage space vector with respect to the direct-axis of the stator reference frame in Figure. 4.2, and $\overline{i_s} = |i_s| e^{j\alpha_s}$.

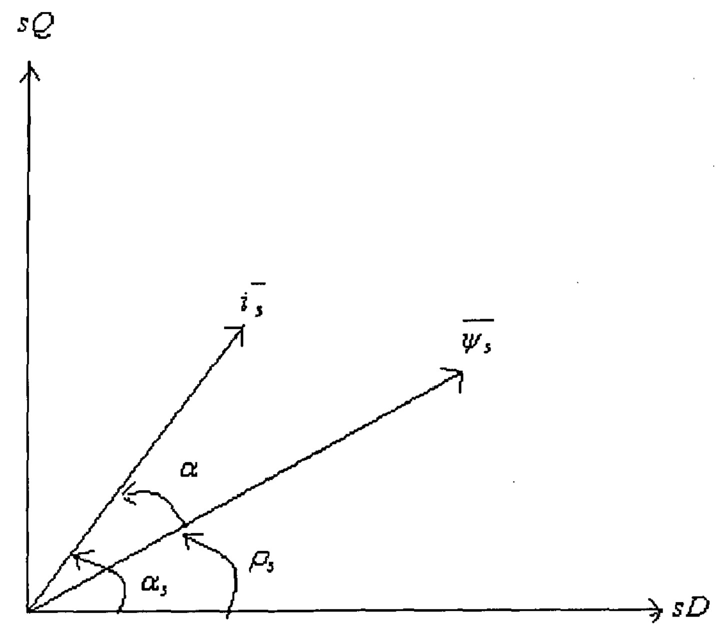

Figure 4.2 Stator flux linkage and stator current space vector [7].

The electromagnetic torque equation can be modified and expressed as follows

$$t_e = \frac{3}{2} P |\psi_s| . |i_s| . \sin(\alpha_s - \rho_s) = \frac{3}{2} P |\psi_s| . |i_s| . \sin(\alpha)$$

Where ρ_s is the stator flux angle, α_s is the stator current, $\alpha = \alpha_s - \rho_s$ is the angle between the stator flux linkage and stator current space vector.

For a given value of the rotor speed, if the modulus of the stator flux linkage space vector is kept constant and the angle ρ_s is changed quickly, then the electromagnetic torque can be rapidly changed.

For mathematical proof, the electromagnetic torque response of the machine for a step change in ρ_s at t=0. The rotor-current space vector (formulated in the stationary reference frame) is expressed in terms of the stator flux-linkage space vector, and also the rotor flux linkage space vector is expressed in terms of the stator flux-linkage space vector.

$$\overline{i_r'} = \left(\overline{\psi_s} - L_s\overline{i_s}\right)\Big/L_m \quad , \quad \overline{\psi_r'} = L_r\overline{i_r'} + L_m\overline{i_s}$$

$$\text{then } \overline{\psi_r'} = \left(L_r\big/L_m\right)\left(\overline{\psi_s} - L_s'\overline{i_s}\right)$$

The expressions for $\overline{i_r'}$ and $\overline{\psi_r'}$ are substituted into the rotor voltage equation expressed in the stationary reference frame as

$$0 = R_r\overline{i_r'} + \frac{d\overline{\psi_r'}}{dt} - j\omega_r\overline{\psi_r'}$$

The derived voltage equation contains $\overline{i_s}$ and $\overline{\psi_s}$, and this can be used to express the stator-current space vector in terms of the stator flux-linkage space vector. This expression for the stator-current space vector is then substituted into the electromagnetic torque equation. However, by also utilizing in this expression the fact that the stator flux linkage space vector modulus is constant.

$$\overline{\psi_s} = \left|\overline{\psi_s}\right|\exp(j\rho_s) = c_1\exp(j\rho_s) \quad \text{and} \quad d\overline{\psi_s}\big/dt = j\left|\overline{\psi_s}\right| d\rho_s\big/dt$$

Finally it is possible to obtain an equation for the electromagnetic torque whose inverse Laplace transform gives the required temporal variation of the electromagnetic torque. An examination of this expression shows that for constant $\left|\overline{\psi_s}\right|$, the rate of change of the increasing electromagnetic torque is almost proportional to the rate of change of ρ_s. Thus by forcing the largest under the condition of constant stator flux-linkage modulus, the fastest electromagnetic torque response time is obtained.

If stator voltages are imposed on the motor, which keep the stator flux constant (at the demanded value), but which quickly rotate the stator flux-linkage space vector into the position required (by the torque demand), then fast torque control is performed. It follows that if in the DTC induction motor drive, the developed actual electromagnetic torque of the machine is smaller than its reference value, the electromagnetic torque should be increased as fast as possible by using the fastest $d\rho_s/dt$. However, when the electromagnetic torque is equal to its reference value, the rotation is stopped. If the stator flux-linkage space vector is accelerated in the forward direction, then positive electromagnetic torque is produced, and when it is decelerated backwards, negative electromagnetic torque is produced. However, the stator flux-linkage vector can be adjusted by using the appropriate stator-voltage space vector, which is generated by the VSI which supplies the induction machine.

For simplicity it is assumed that the stator ohmic drops can be neglected, then $d\overline{\psi_s}/dt = \overline{u_s}$, and it can be seen that the inverter voltage ($\overline{u_s} = \overline{u_i}$) directly impresses the stator flux, and thus the required stator-flux locus will be obtained by using the appropriate inverter voltages obtained by using the appropriate inverter switching states. In a short time, when the voltage vector is applied, $\Delta\overline{\psi_s} = \overline{u_s}\Delta t$. Thus the stator flux-linkage space vector moves by $\Delta\overline{\psi_s}$ in the direction of the stator voltage space vector at a speed proportional to the magnitude of the stator voltage space vector. By selecting step-by-step the appropriate stator voltage vector, it is then possible to change the stator flux in the required way [7,26]. Decoupled control of the torque and stator flux is achieved by acting on the radial and tangential components of the stator flux-linkage space vector locus. These two components are directly proportional to the components of the stator voltage space vector in the same directions and thus can be controlled by the appropriate inverter switching's.

The electromagnetic torque can be quickly changed by controlling the stator flux-linkage space vector, which, however, can be changed by using the appropriate stator voltages. It can be seen that there is direct stator-flux and electromagnetic torque control achieved by using the appropriate stator voltages. This is why this type of control is usually referred to as direct torque control.

4.3 Basic Direct Torque Control Strategy

The control block diagram of the stator flux based DTC induction motor drive with VSI is shown in Figure. 4.3. The control block is composed of Flux comparator, Torque comparator, Inverter optimum switching table, Electromagnetic torque and Stator flux-linkage estimator and Voltage source inverter [29,30].

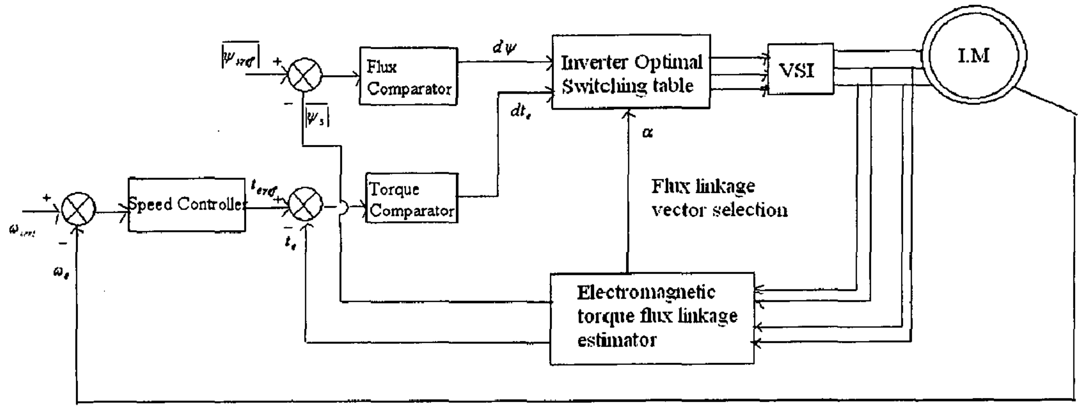

Figure 4.3 Schematic of stator flux based DTC induction motor drive

The MATLAB model of the DTC is shown below:

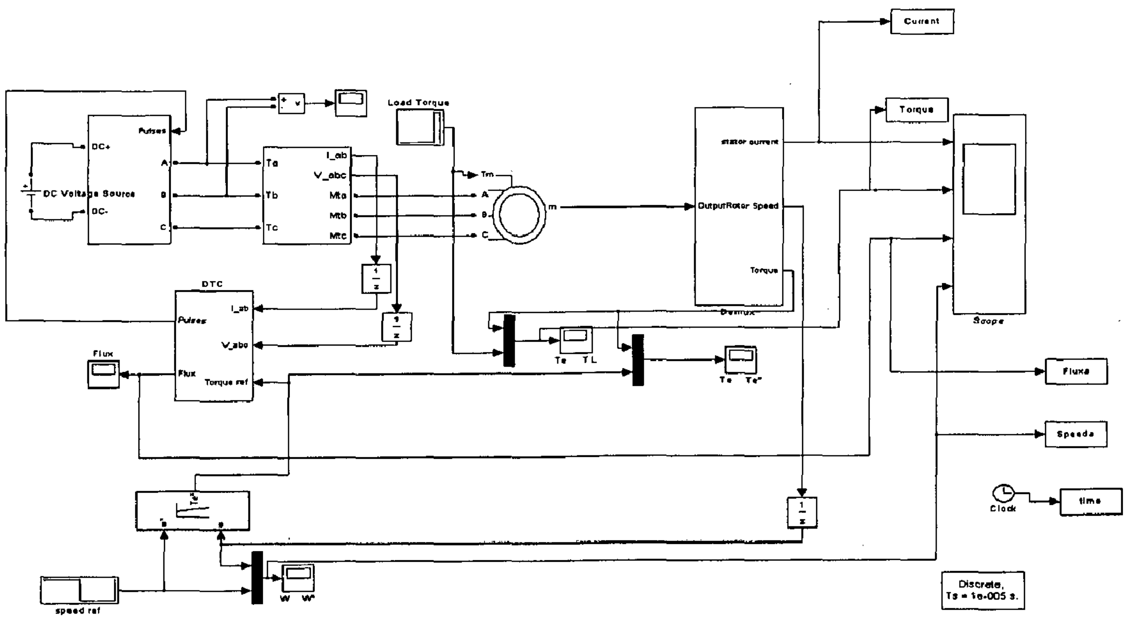

Figure 4.4 MATLAB model of DTC induction motor drive

4.3.1 Stator Flux-linkage Space Vector Position Estimation

The position of the stator flux-linkage space vector that means stator flux angle (ρ_s) can be determined by using the estimated values of the direct and quadrature-axis stator flux-linkages in the stationary reference frame (ψ_{sD} , ψ_{sQ}), thus

The stator flux angle (ρ_s) is

$$\rho_s = \tan^{-1}(\psi_{sQ} / \psi_{sD})$$

4.3.2 Electromagnetic Torque and Stator Flux-linkage Estimator

The stator flux linkage space vector is

$$\overline{\psi_s} = \int \left(\overline{u_s} - R_s \overline{i_s} \right) dt$$

Where $\overline{\psi_s} = \psi_{sD} + j\psi_{sQ}$; $\overline{u_s} = u_{sD} + ju_{sQ}$ and $\overline{i_s} = i_{sD} + ji_{sQ}$

$$\psi_{sD} = \int \left(u_{sD} - R_s i_{sD} \right) dt$$

$$\psi_{sQ} = \int \left(u_{sQ} - R_s i_{sQ} \right) dt$$

$$u_{sD} = \frac{1}{3}\left(u_{AB} - u_{CA}\right), \quad u_{sQ} = \frac{1}{\sqrt{3}}\left(u_{AC} + u_{BA}\right) \quad i_{sD} = \frac{2}{3}\left(i_{as} - \frac{1}{2}i_{bs} - \frac{1}{2}i_{cs}\right), \quad i_{sQ} = \frac{1}{\sqrt{3}}\left(i_{bs} - i_{cs}\right)$$

Where $\psi_{sD}, u_{sD}, i_{sD}$ and ψ_{sQ}, u_{sQ}, i_{sQ} are the d-axis and q-axis quantities of stator flux-linkage space vector, stator voltage space vector and stator current space vector components respectively.

4.3.3 Flux and Torque Hysteresis Comparators

The error between the estimated torque and the reference torque is the input of a three level hysteresis comparator, whereas the error between the estimated stator flux magnitude and the reference stator flux magnitude is the input of a two level hysteresis comparator [32,33].

The selection of the appropriate voltage vector is based on the switching table given in Table 4.1. The input quantities are the stator flux sector and the outputs of the two hysteresis comparators. Figure 4.5 a) and 4.5 b) illustrate the torque and flux comparators, respectively [32].

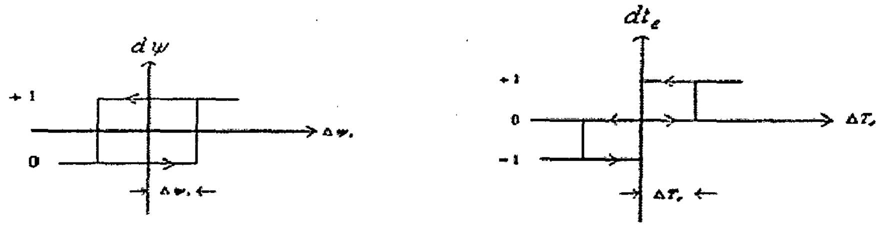

Figure 4.5 a) Flux Hysteresis Comparator b) Torque Hysteresis Comparator

In Figure 4.6 in the MATLAB model shown, the reference value of the stator flux-linkage space vector modulus, $|\overline{\psi_{sref}}|$, is compared with the actual modulus of the stator flux-linkage space vector, $|\overline{\psi_s}|$ and the resulting error is fed into the two-level stator flux hysteresis comparator. Similarly the reference value of the electromagnetic torque, t_{eref}, is compared with the actual electromagnetic torque t_e, and the resulting torque error is fed into the three-level torque hysteresis comparator.

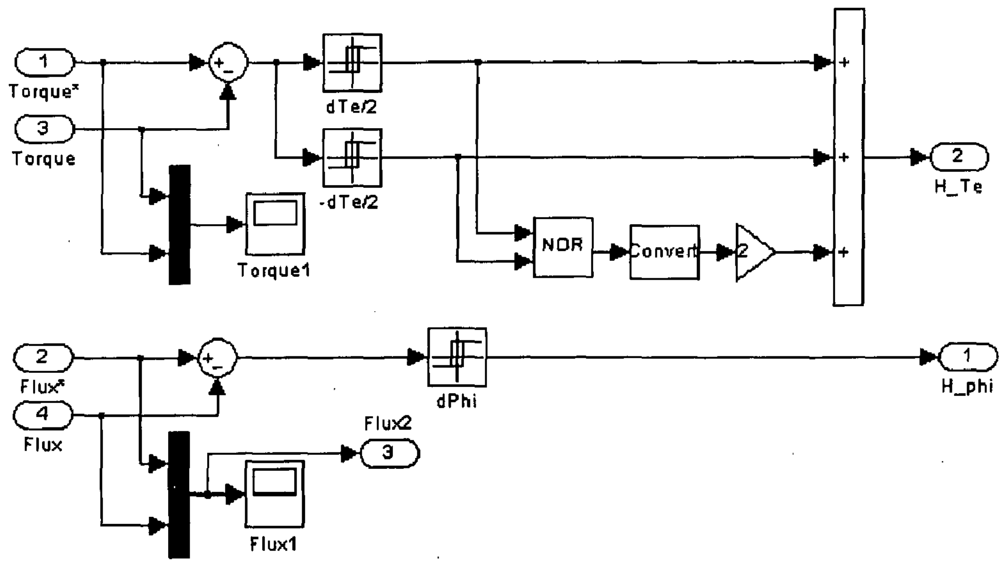

Figure 4.6 MATLAB model of Flux and Torque Comparators

4.4 Voltage Source Inverter

The Voltage source inverter (VSI) is a six-pulse inverter. The VSI getting control pulses from the Inverter optimal switching table. The stator flux-linkage space vector is

$$\overline{\psi_s} = \int \left(\overline{u_s} - R_s \overline{i_s} \right) dt$$

From equation, if the stator ohmic drops can be neglected, then the equation can be modified as shown below $d\overline{\psi_s}/dt = \overline{u_s}$

From the equation, $d\overline{\psi_s}/dt = \overline{u_s}$, in a short Δt time, when the voltage vector is applied, $\overline{\Delta\psi_s} = \overline{u_s}\Delta t$, that means the stator flux-linkage moves by $\overline{\Delta\psi_s}$ in the direction of stator-voltage space vector at a speed which proportional to the magnitude of the stator-voltage space vector.

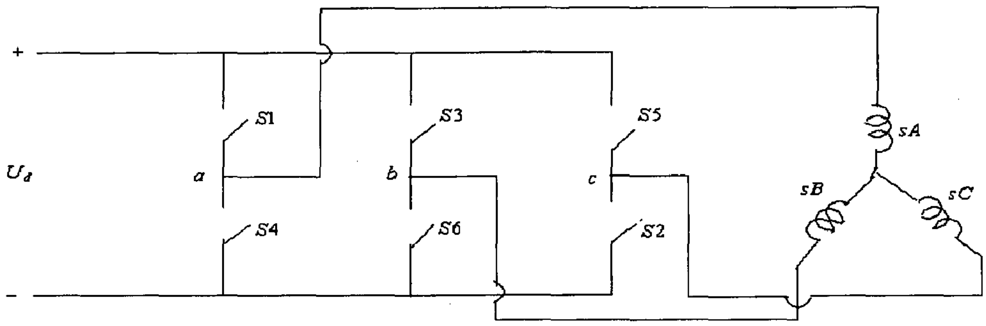

Figure 4.7 Voltage Source Inverter [7]

By considering the six-pulse VSI shown in Figure 4.7, there are six non zero active voltage-switching space vectors ($\overline{u_1}, \overline{u_2}, \overline{u_3}, \overline{u_4}, \overline{u_5}, \overline{u_6}$) and two zero space vectors ($\overline{u_7}, \overline{u_8}$). These are shown in Appendix C. The six active inverter-switching vectors can be expressed as

$$\overline{u_s} = \overline{u_k} = \frac{2}{3} U_d \exp\left[\frac{j(k-1)\pi}{3}\right] \qquad \text{k=1, 2, 3, 4, 5, 6}$$

Where U_d is the d.c. link voltage. For k=7, 8, $\overline{u_k} = 0$ holds for the two zero switching states where the stator windings are short circuited $\overline{u_s} = \overline{u_k} = 0$.

4.5 Stator Flux-linkage Space Vector Control

The stator flux linkage space vector will move fast if non-zero switching vectors are applied, for a zero vector it will almost stop (it will move very slowly due to the small ohmic voltage drop). In the DTC drive, at every sampling period, the switching vectors are selected such that the stator flux-linkage error and torque errors should be confined within the respective hysteresis bands. The drive may lose the control, if the width of the hysteresis bands is too small value. It is assumed that the widths of these hysteresis bands are $2\,\Delta\overline{\psi_s}$ and $2\,\Delta t_e$ respectively.

If the selected switching vector is zero, then the speed of the stator flux-linkage space vector is zero, it is possible to change this speed by changing the output ratio between the zero and non-zero voltage vectors and the duration of the zero states has a direct effect on the electromagnetic torque oscillations [7]. If a reduced stator flux-linkage space vector is required, it can be achieved by applying switching voltage vectors which are directed towards the centre of the rotor, and if an increased stator flux-linkage space vector is required, it can be achieved by applying switching voltage vectors which are directed out from the centre of the rotor, which is shown in Figure 4.8.

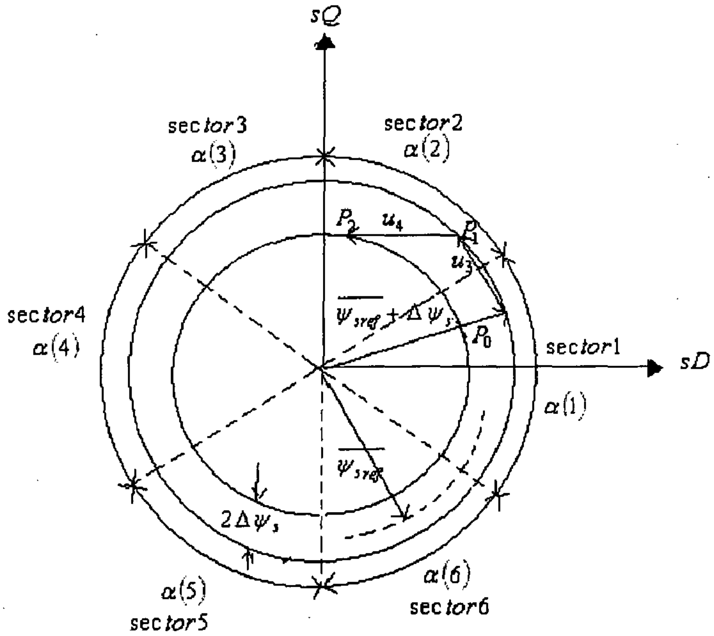

Figure 4.8 Control of the stator flux linkage space vector: stator flux linkage space vector locus [stator flux variations $\Delta\psi_s$] and inverter switching vectors [7]

To keep the modulus of the stator flux linkage space vector ($|\overline{\psi_s}|$) within the hysteresis band, whose width is $2\Delta\overline{\psi_s}$ as shown in Figure. 4.8. The locus of the flux linkage space vector is divided into several sectors, and due to the six step inverter, the minimum number of sectors required is six. The sectors are also shown in Figure as dotted lines.

There are eight switching vectors to keep the modulus of the stator flux linkage space vector ($|\overline{\psi_s}|$) within the hysteresis band. It is assumed that initially the stator flux-linkage space vector is at position P_0, thus in sector1. Assuming that the stator flux-linkage space vector is rotating in anticlockwise, from Figure, by observing the stator flux-linkage space vector at position P_0, it is at the upper limit ($|\overline{\psi_{sref}}|+\Delta\overline{\psi_s}$) so it must be reduced. This can be achieved by applying the suitable switching vector $\overline{u_3}$. Thus the stator flux-linkage space vector will move rapidly from point P_0 to the point P_1 which is in the sector 2. At point P_1 the stator flux-linkage space vector is again at its upper limit, it has to be reduced when it is rotated in anti clockwise, hence for this purpose the switching vector $\overline{u_4}$ has to be selected, and then $\overline{\psi_s}$ moves from point P_1 to point P_2 as shown in Figure 4.8.

On the other hand, if the stator flux-linkage space vector moves in the clockwise direction from point P_0, then the switching vector $\overline{u_5}$ would have to be selected, since this would ensure the required rotation and also the required flux decrease. On the other hand, if at point P_1 the rotation of the stator flux-linkage space vector has to be stopped, then a zero switching vector would have to be applied, so either $\overline{u_7}$ or $\overline{u_8}$ can be applied. However, since prior to this the last switching was performed by the application of the switching vector $\overline{u_3}=\overline{u_3}(010)$, which means that the first switch is connected to the (lower) negative d.c. rail, the second switch is connected to the (upper) positive d.c. rail, and the third switch is connected to the negative d.c. rail, to minimize the number of switching's, the state $\overline{u_8}(000)$ is selected, since it requires only switching of the second switch from 1 to 0 in contrast too selecting $\overline{u_7}(111)$, which would require two switching's.

If the stator flux-linkage space vector is at point P_2, then the lower limit $\left(\left|\overline{\psi_{sref}}\right| - \Delta\psi_s\right)$ is reached and the stator flux-linkage space vector can be rotated in the anti-clockwise direction to point P_3 by increasing it, and for this purpose the switching vector $\overline{u_3}$ gives the fastest rotation. It can be seen that point P_3 is still in sector 2. If on the other hand, the flux-linkage space vector has to be rotated from P_2 in the opposite direction then by selecting the switching vector which rotates $\overline{\psi_s}$ from P_2 (where the flux-linkage space vector is at the lower limit, thus flux increase is required) in the fastest way in the clockwise direction gives the switching vector $\overline{u_1}$, etc.

When the electromagnetic torque has to be changed then the stator flux-linkage space vector has to be rotated in the appropriate direction. If an increase of the torque is required, then the torque is controlled by applying voltage vectors that advance the flux linkage space vector in the direction of rotation and if a decrease is required, voltage vectors are applied which oppose the direction of the torque and If zero torque is required then zero switching vector is applied ($\overline{u_7}$ or $\overline{u_8}$), which minimizes the inverter switching. It follows that the angle of the stator voltage space vector is indirectly controlled through the flux vector modulus and torque, and increasing torque causes an increased torque angle.

For example, when the stator flux rotates anticlockwise, and if an increase in the electromagnetic torque is required, then if the stator flux-linkage space vector is in the second sector at point P_1 where the flux linkage has to be decreased, then the electromagnetic torque increase can be achieved by applying switching vector $\overline{u_4}$.

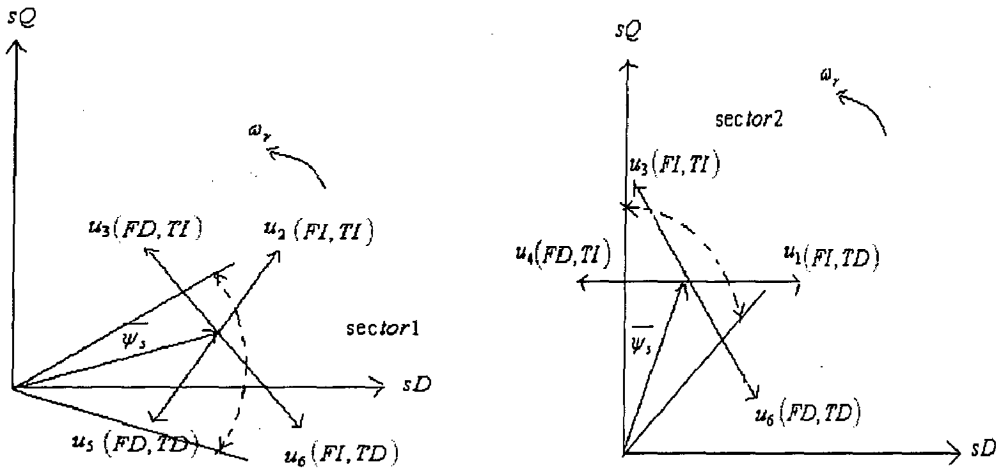

Figure 4.9 Position of various stator flux-linkage space vectors, and selection of the optimum switching voltage vectors. FI: flux increase; FD: flux decrease; TI: torque increase; TD: torque decrease [7].

On the other hand if the stator flux-linkage space vector is in the second sector but a torque decrease is required, but the flux-linkage has to be increased, then this can be achieved by applying the switching vector $\overline{u_1}$, since this moves the stator flux-linkage space vector in the clockwise direction (which is the direction for the negative torque), and also increases the stator flux linkage. If the stator flux-linkage space vector is in the second sector and a torque decrease is required, but the stator flux linkage has to be decreased, then the switching vector $\overline{u_6}$ has to be applied, etc. Figure shows the position of the various stator flux linkage vectors if the stator flux linkage space vector is in one of the six sectors. It is also shown which switching vector has to be selected to obtain the required increase or decrease of the stator flux linkage and the required increase or decrease of the electromagnetic torque.

4.6 Inverter Optimal Switching Table

The outputs of the two comparators $(d\psi, dt_e)$ are used in the inverter optimal switching table (look-up-table), which also uses the information on the position of the stator flux-linkage space vector. The table is known as optimum switching vector selection table. This gives the optimum selection of the switching vectors for all the possible stator flux-linkage vector positions.

The digital output signals of a two level flux hysteresis comparator are

$d\psi=1$ if $|\overline{\psi}_s| \leq |\psi_{sref}| - |\Delta\psi_s|$, $d\psi=1$ if a stator flux increase is required.

$d\psi=0$ if $|\overline{\psi}_s| \geq |\overline{\psi}_{sref}| + |\Delta\psi_s|$, $d\psi=1$ if a stator flux decrease is required.

The digital output signals of a three level torque hysteresis comparator are

- For the stator flux-linkage space vector rotates in the forward direction (anti clockwise direction)

  $dt_e=1$ if $|t_e| \leq |t_{eref}| - |\Delta t_e|$ if a torque increase is required

  $dt_e=0$ if $t_e \geq t_{eref}$ if no change in the torque is required

- For the stator flux-linkage space vector rotates in the reverse direction (clockwise direction)

  $dt_e=-1$ if $|t_e| \geq |t_{eref}| + |\Delta t_e|$ if a torque decrease is required

  $dt_e=0$ if $t_e \leq t_{eref}$ if no change in the torque is required

$d\psi$	dt_e	$\alpha(1)$ sector1	$\alpha(2)$ Sector2	$\alpha(3)$ sector3	$\alpha(4)$ sector4	$\alpha(5)$ sector5	$\alpha(6)$ sector6
	1	$\overline{u}_2$	$\overline{u}_3$	$\overline{u}_4$	$\overline{u}_5$	$\overline{u}_6$	$\overline{u}_1$
1	0	$\overline{u}_7$	$\overline{u}_8$	$\overline{u}_7$	$\overline{u}_8$	$\overline{u}_7$	$\overline{u}_8$
	-1	$\overline{u}_6$	$\overline{u}_1$	$\overline{u}_2$	$\overline{u}_3$	$\overline{u}_4$	$\overline{u}_5$
	1	$\overline{u}_3$	$\overline{u}_4$	$\overline{u}_5$	$\overline{u}_6$	$\overline{u}_1$	$\overline{u}_2$
0	0	$\overline{u}_8$	$\overline{u}_7$	$\overline{u}_8$	$\overline{u}_7$	$\overline{u}_8$	$\overline{u}_7$
	-1	$\overline{u}_5$	$\overline{u}_6$	$\overline{u}_1$	$\overline{u}_2$	$\overline{u}_3$	$\overline{u}_4$

Table 4.1 Optimum voltage switching vector look up table [32]

4.7 RESULTS

The simulation has been performed using actual parameters of 1 HP and 30 HP induction motor fed by an IGBT inverter for starting, speed reversal, speed re-reversal and load perturbation and the results have been shown in Figure 4.10 and Figure 4.11resp.

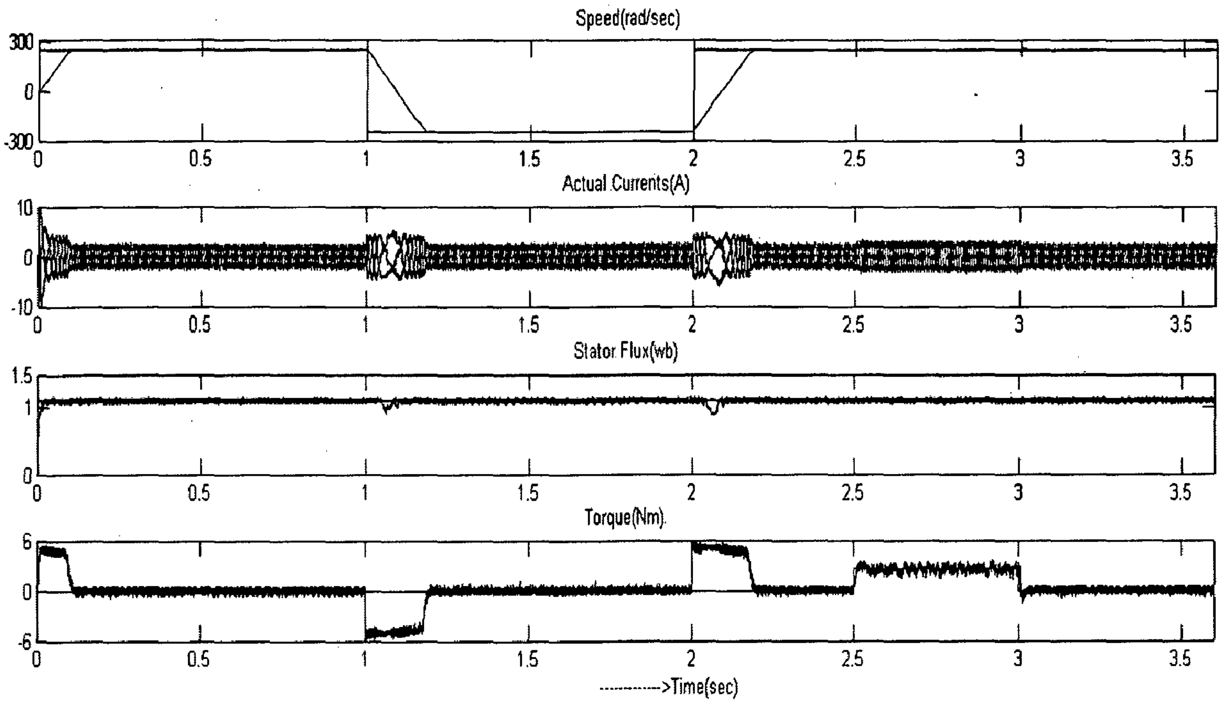

Figure 4.10 Starting, Speed Reversal, Speed Re-reversal and Load Perturbation Response of 1HP Direct Torque Controlled Induction Motor Drive system.

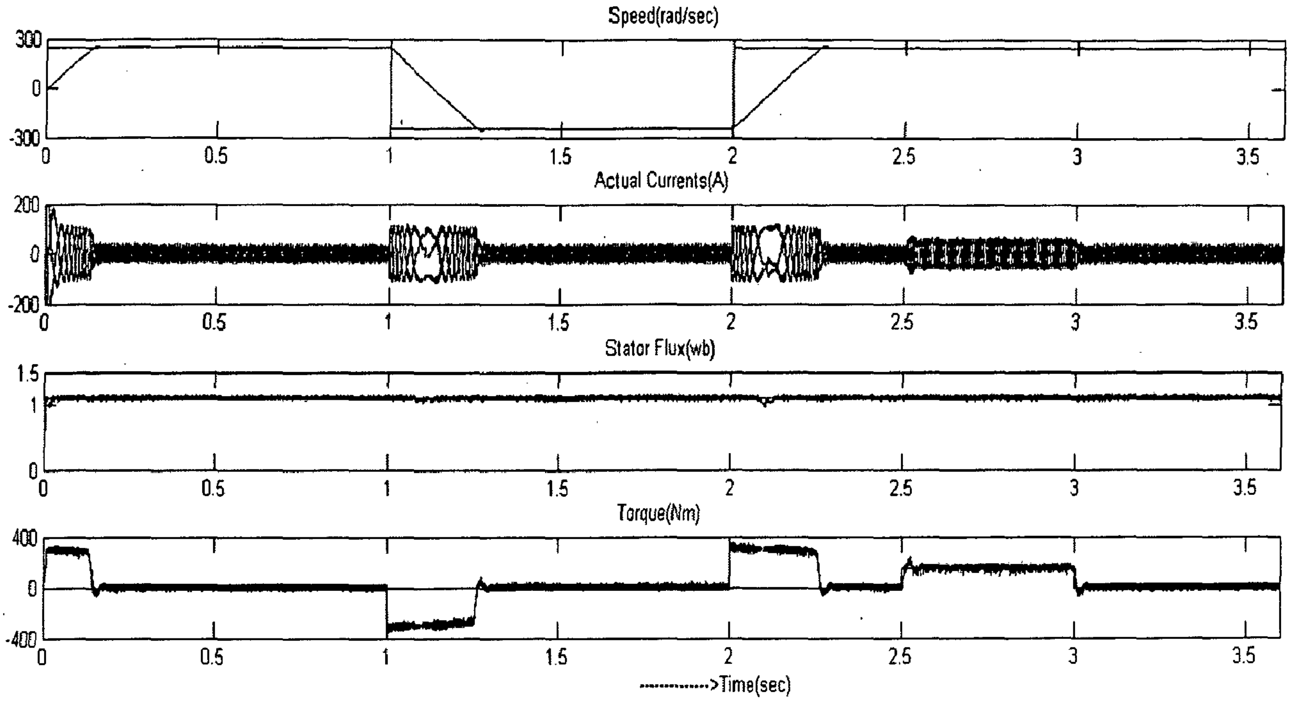

Figure 4.11 Starting, Speed Reversal, Speed Re-reversal and Load Perturbation Response of 30HP Direct Torque Controlled Induction Motor Drive system.

4.8 CONCLUSION

In this chapter, the basic concept of the direct torque control technique for the induction motor drive has been discussed and the MATLAB model of the technique is explained. The estimation of the electromagnetic torque and the stator flux-linkage is presented. The explanation for the hysteresis comparators of the torque and flux is provided. The selection of the voltage vectors for various demands such as flux increase, decrease and torque increase, decrease are explained thoroughly and at last the switching table for the selection of the voltage vectors depending on the condition obtained from the hysteresis comparator is presented.

CHAPTER V

COMPARISON OF VECTOR CONTROL AND DIRECT TORQUE CONTROL

5.0 INTRODUCTION

The performance of vector control and direct torque control modes of operation of induction motor drive system is simulated in MATLAB environment using simulink and power system blockset toolboxes under different dynamic conditions such as starting, speed reversal and load perturbation(load application and load removal) has been observed. The simulation is carried out for induction motors of two different ratings namely, 1HP (0.75KW) and 30HP (22KW) respectively. The present chapter provides a comparative study between the two different control techniques considered on the basis of the observations recorded from the simulation process.

5.1 STARTING DYNAMICS

Three phase squirrel cage induction motor is fed from controlled voltage and frequency source. Motor is started at low frequency decided by the controller and finally runs at the steady state condition at the set reference speed value. The reference speed is set at 250 elec rad/sec as the base speed is 314.15 elec rad/sec with a torque limit set at twice the rated value. The following results are observed for starting dynamics for both vector control and direct torque control and for 1 HP and 30 HP.

5.2 REVERSAL DYNAMICS

When the reference speed is changed from 250 rad/sec to -250 rad/sec, the motor tends to run in reverse direction. When the controller observes this change, it first reduces the frequency of the stator currents demonstrating the regenerative braking followed by the phase reversal for starting the motor in reverse direction. As the drive is in the same dynamic state (on no load) just before and after the reversal phenomenon, the steady state values of the inverter currents are observed to be same in either directions of the rotation of the motor. However the phase sequence of these currents in two directions will be different. Speed reversal dynamics with vector control and direct torque control are shown in Figure. 5.5 to 5.8.

5.3 LOAD PERTURBATION DYNAMICS

The study of the performance of vector control and direct torque control induction motor drive under load perturbation (load application and load removal) is really important as the speed of the motor should not change under any load conditions. When the motor is running at steady state speed 250 elec rad/sec a load torque equal to the rated torque is applied on the shaft. Sudden application of load causes an instantaneous fall in speed. In response to the drop in speed the output of the speed controller responds by increasing the reference torque value. Therefore the developed electromagnetic torque of the induction motor increases causing the motor speed to settle to the reference value with the increasing winding currents.

When the motor is operating at steady state on load, sudden load removal causes a small overshoot in the motor speed because of this small overshoot input to the speed controller will become negative and consequently the output of the speed controller i.e., the reference torque reduces. Therefore the electromagnetic torque developed by the motor reduces, this causes reduction in the motor speed and hence the speed decreases and settles to the reference value. After the removal of load the stator currents also reduce to the no load value. Load perturbation dynamics with vector control and direct torque control are shown in Figure. 5.13 to 5.16.

The following results are observed for Starting Dynamics for both vector control and direct torque control and for 1 HP and 30 HP.

For 1HP Motor:

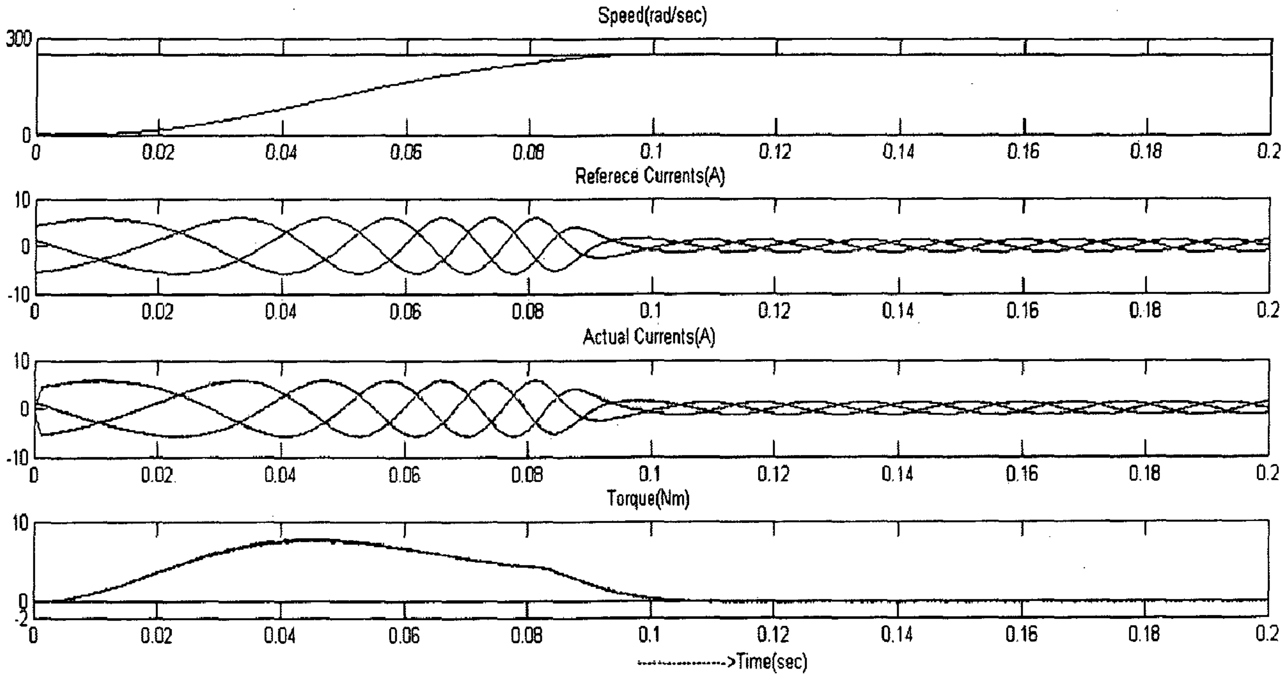

Figure 5.1 Starting Dynamics of 1HP Vector Controlled Induction Motor Drive System

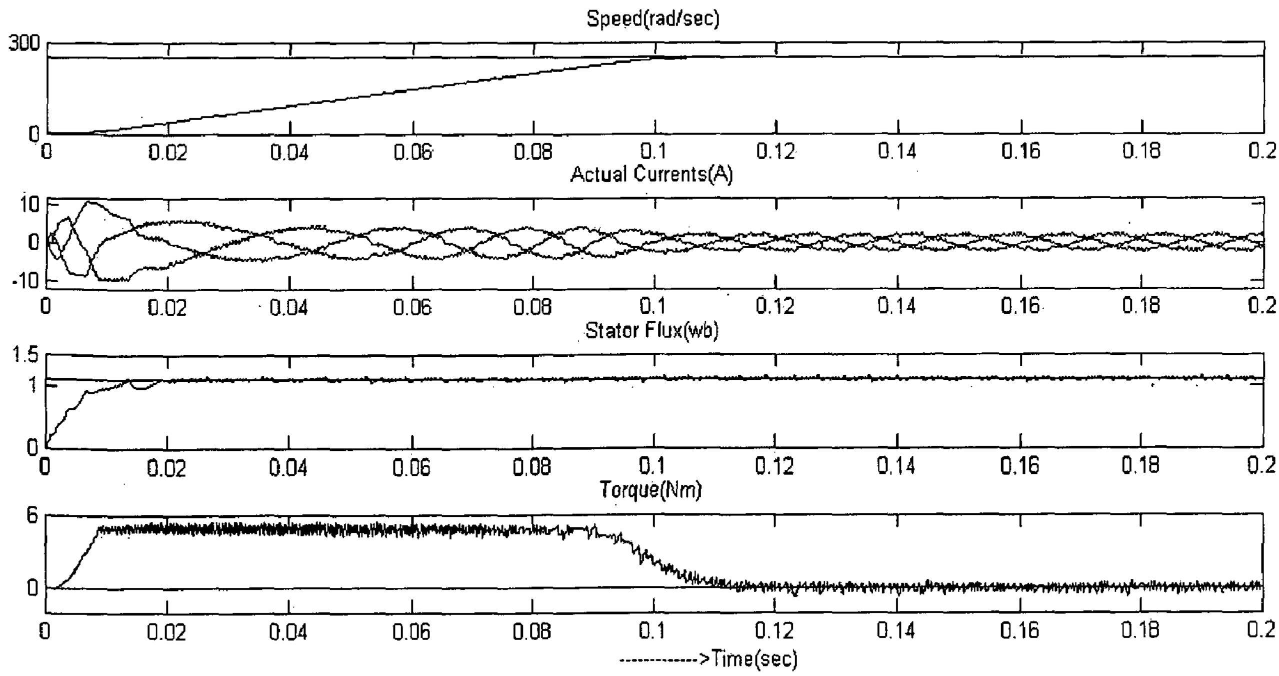

Figure 5.2 Starting Dynamics of 1HP Direct Torque Controlled Induction Motor Drive System

For 30 HP Motor:

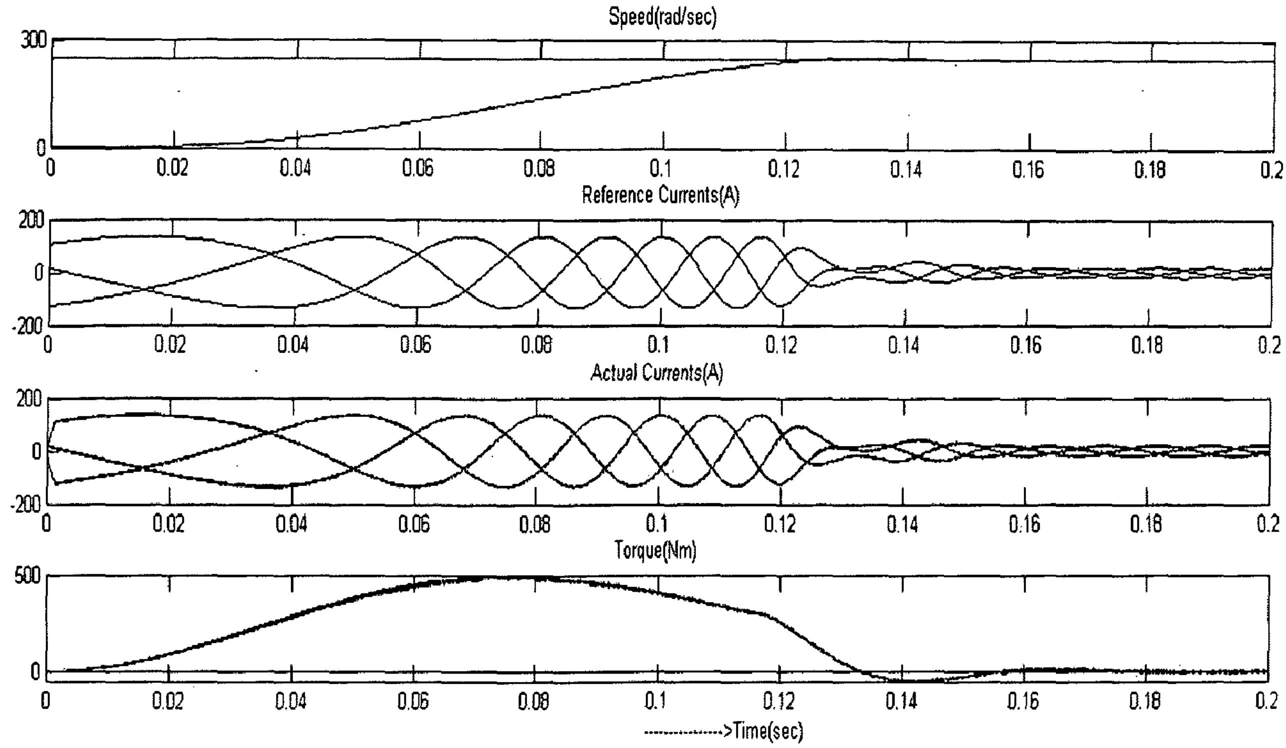

Figure 5.3 Starting Dynamics of 30HP Vector Controlled Induction Motor Drive System

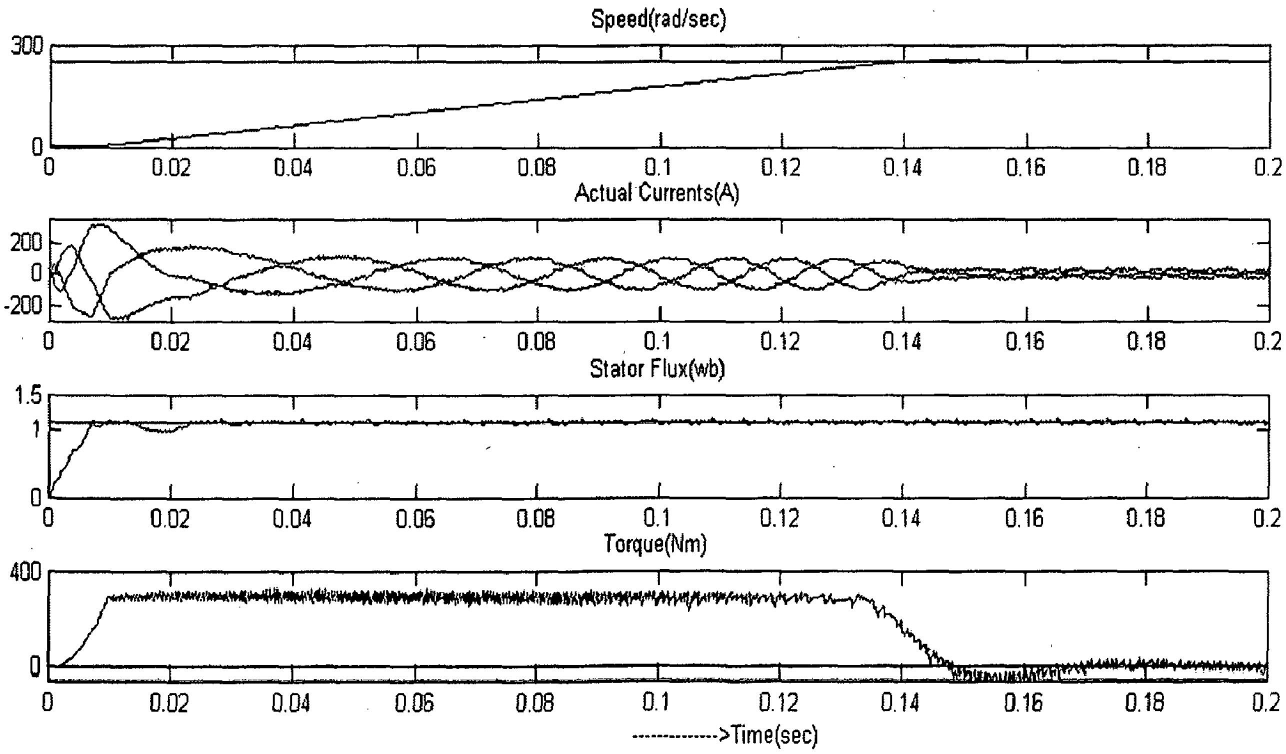

Figure 5.4 Starting Dynamics of 30HP Direct Torque Controlled Induction Motor Drive System

The following results are observed for Reversal Dynamics for both vector control and direct torque control and for 1 HP and 30 HP.

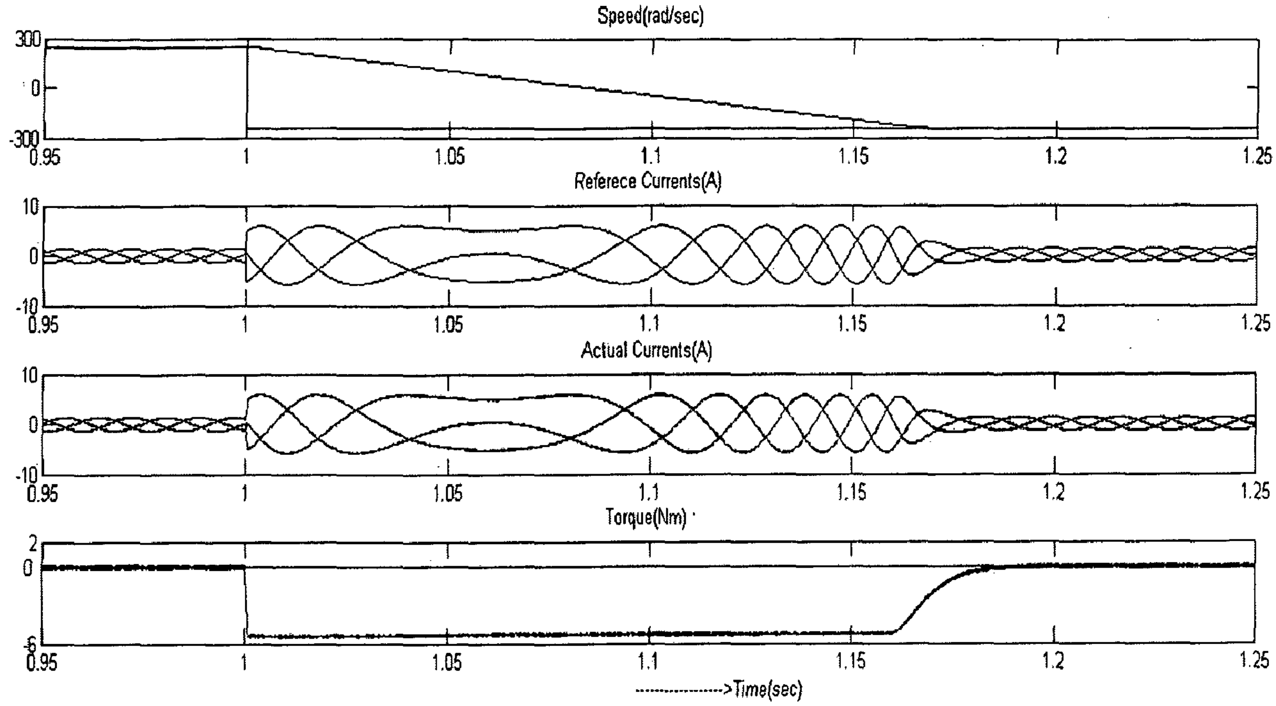

Figure 5.5 Reversal Dynamics of 1HP Vector Controlled Induction Motor Drive System

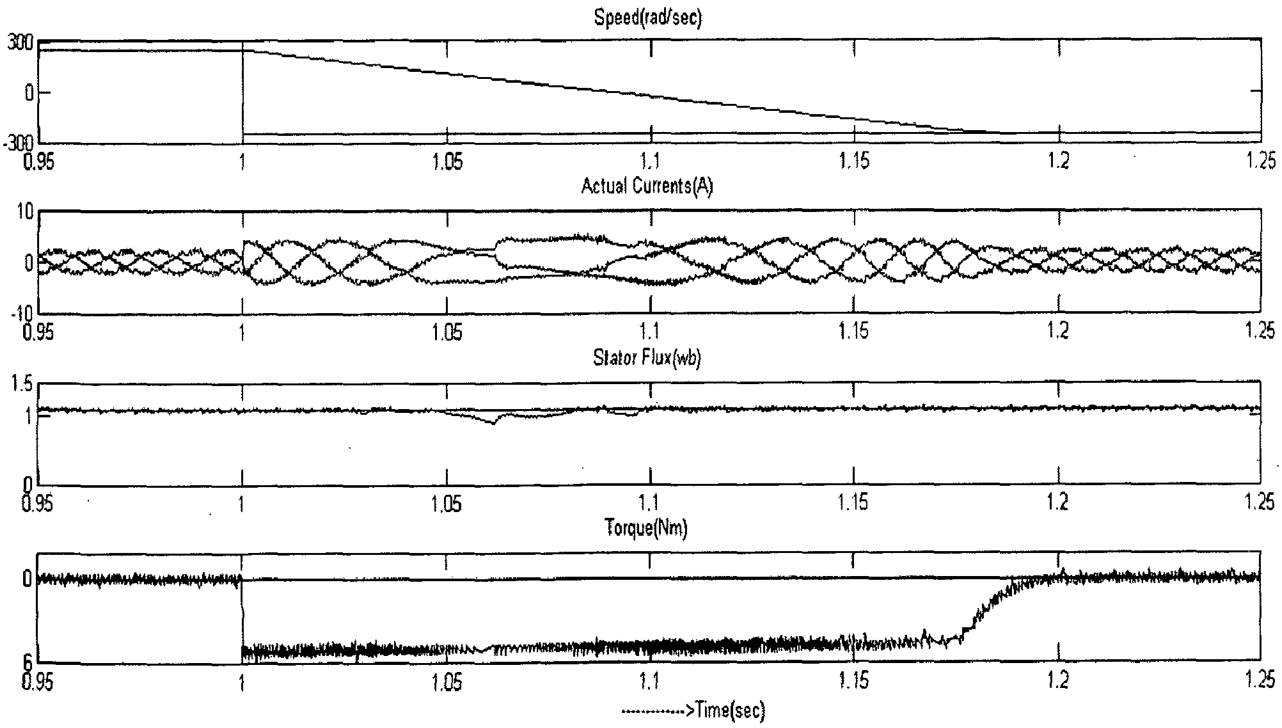

Figure 5.6 Reversal Dynamics of 1HP Direct Torque Controlled Induction Motor Drive System

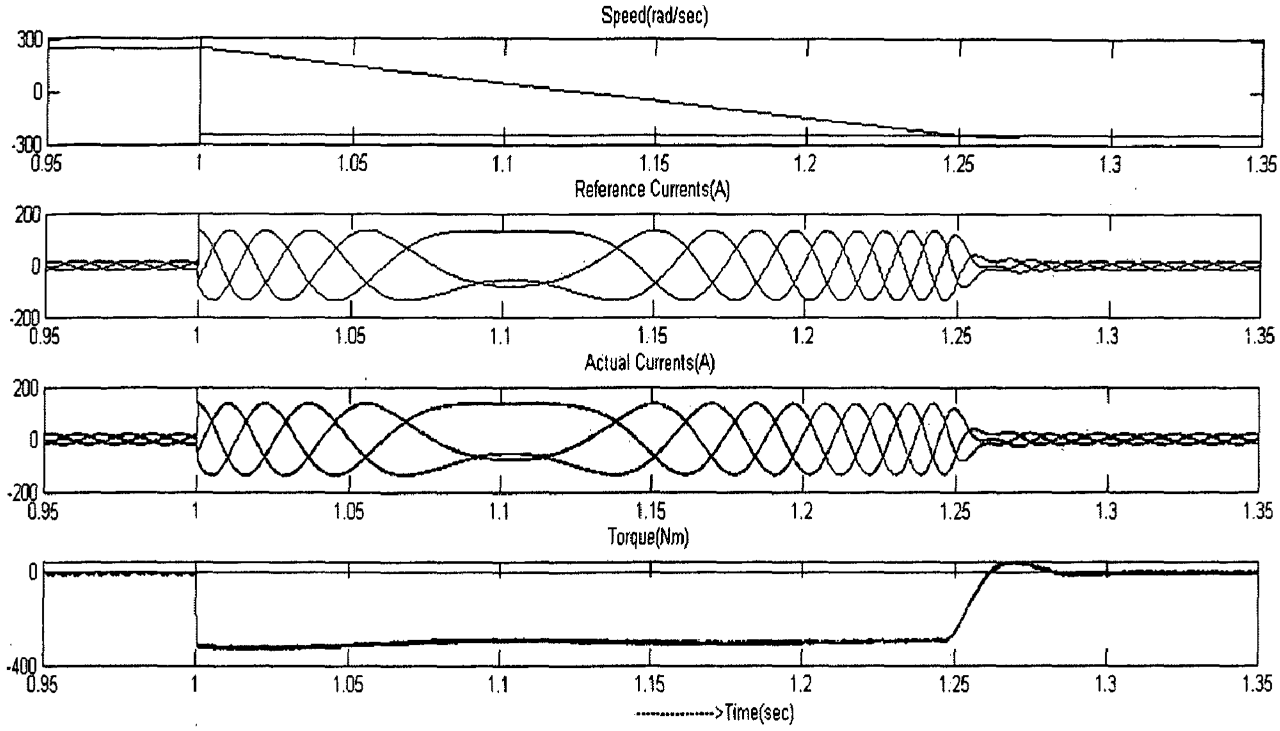

Figure 5.7 Reversal Dynamics of 30HP Vector Controlled Induction Motor Drive System

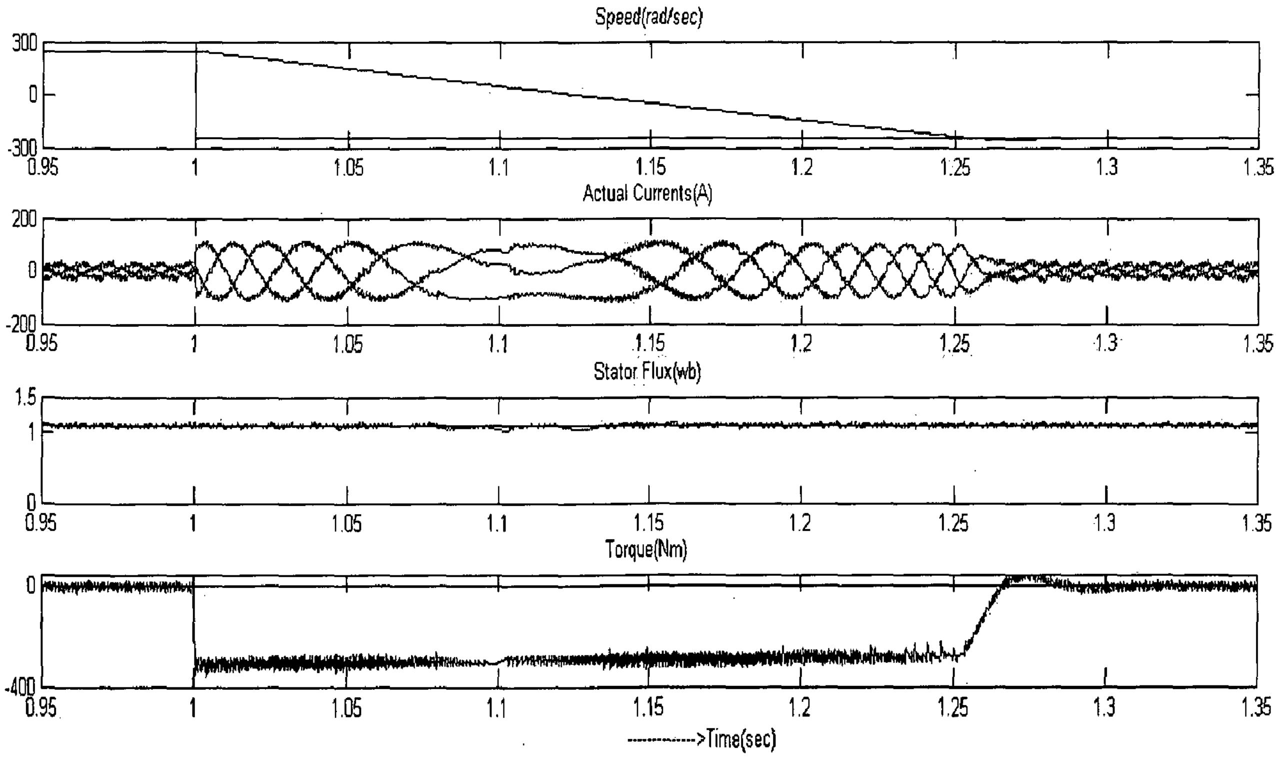

Figure 5.8 Reversal Dynamics of 30HP Direct Torque Controlled Induction Motor Drive System

The following results are observed for Re-Reversal Dynamics for both vector control and direct torque control and for 1 HP and 30 HP.

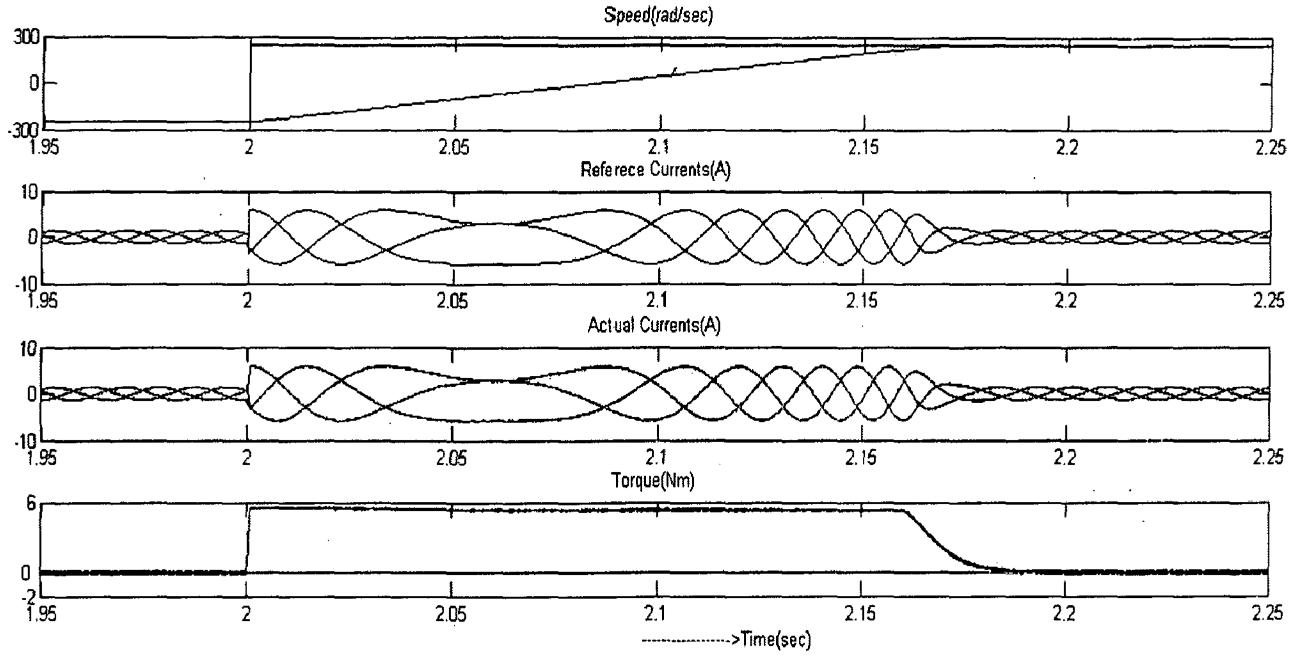

Figure 5.9 Re-Reversal Dynamics of 1HP Vector Controlled Induction Motor Drive System

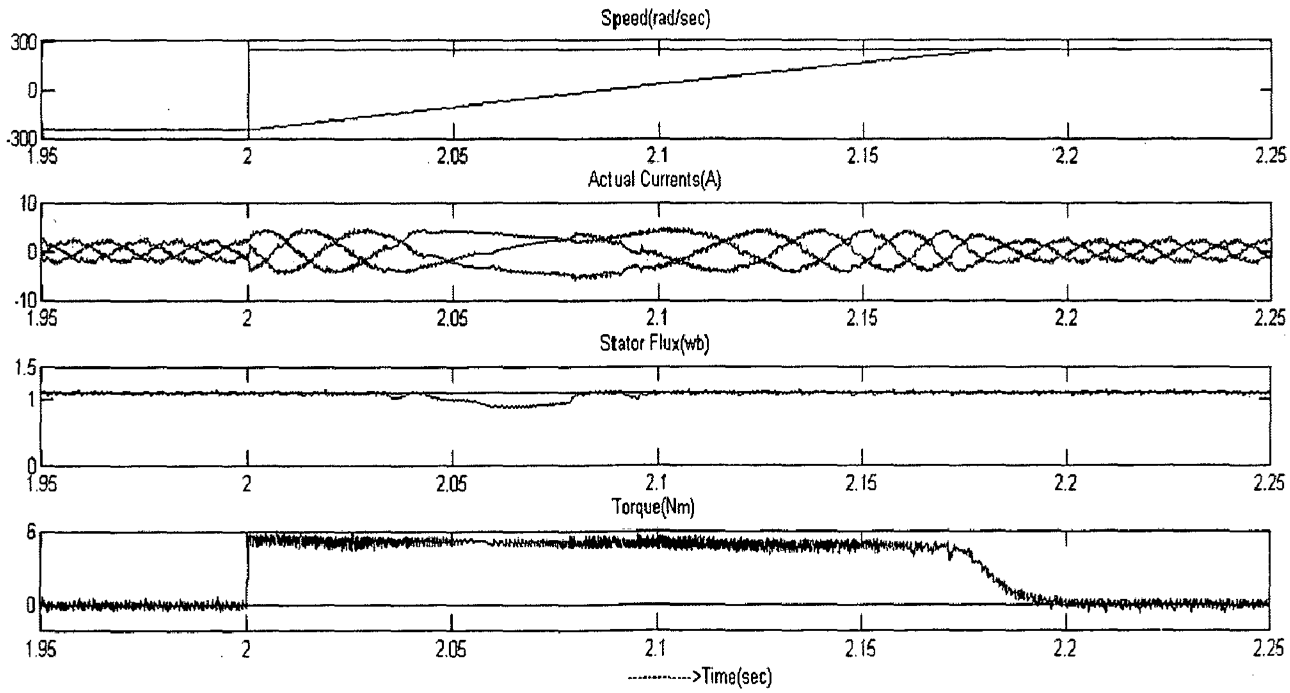

Figure 5.10 Re-Reversal Dynamics of 1HP Direct Torque Controlled Induction Motor Drive System

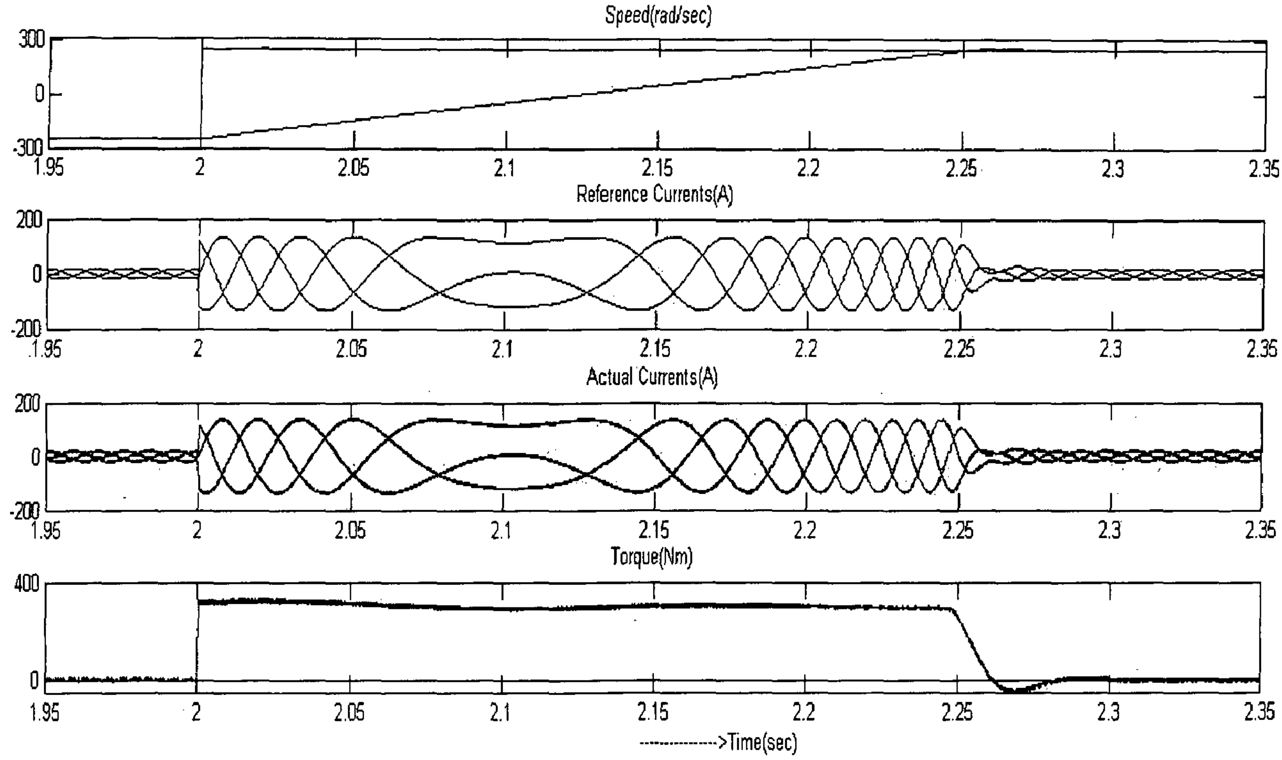

Figure 5.11 Re-Reversal Dynamics of 30HP Vector Controlled Induction Motor Drive System

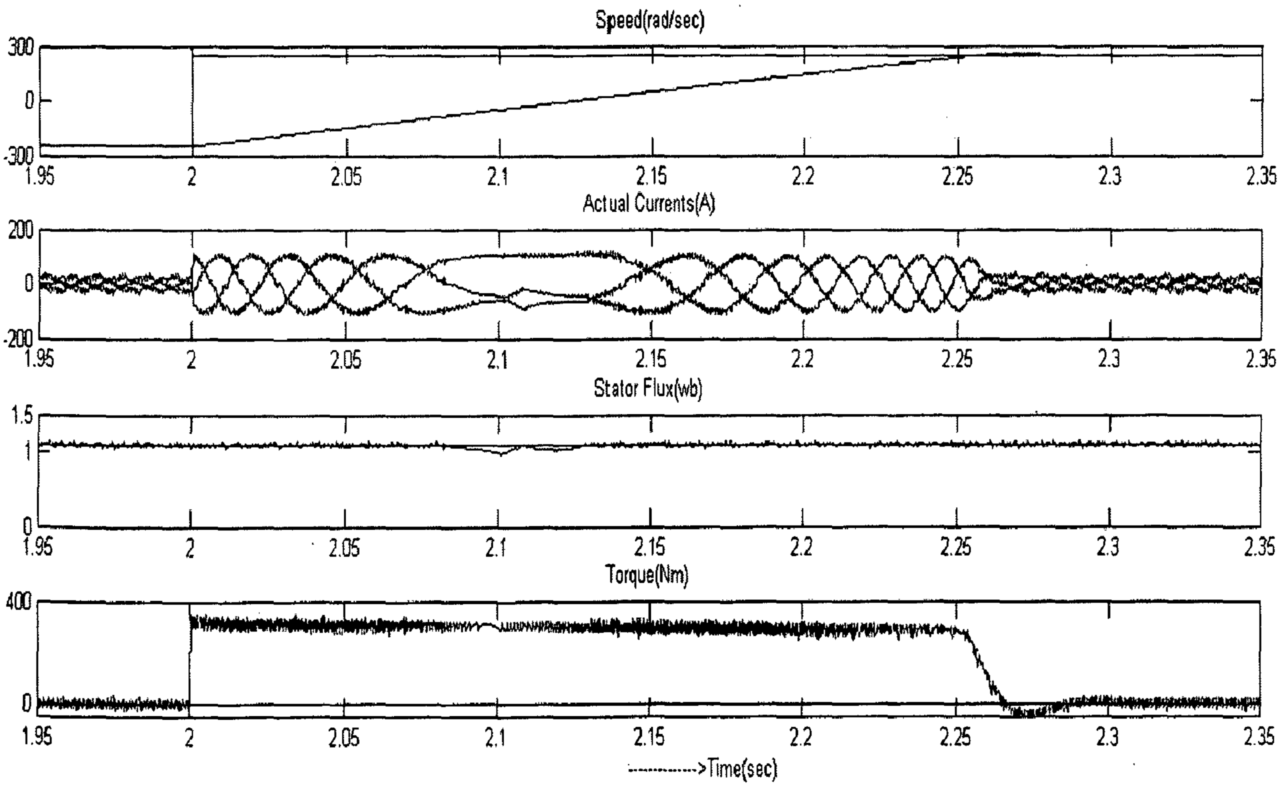

Figure 5.12 Re-Reversal Dynamics of 30HP Direct Torque Controlled Induction Motor Drive System

The following results are observed for Load Perturbation Dynamics for both vector control and direct torque control and for 1 HP and 30 HP.

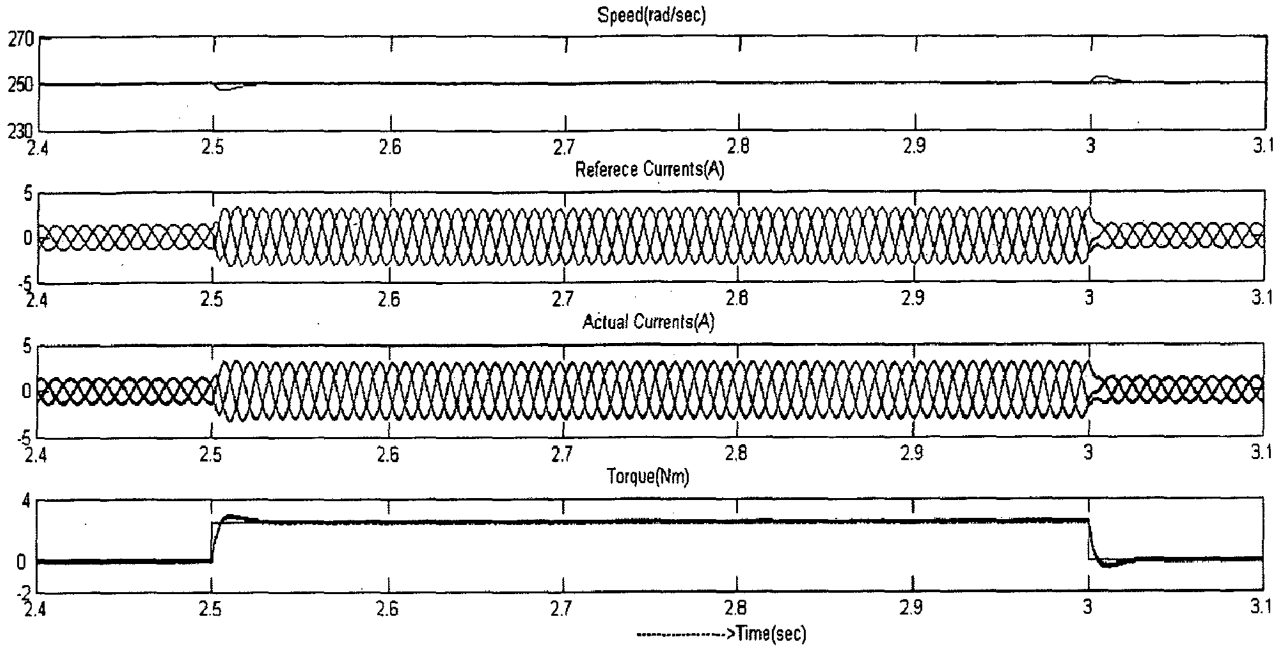

Figure 5.13 Load Perturbation Dynamics of 1HP Vector Controlled Induction Motor Drive System

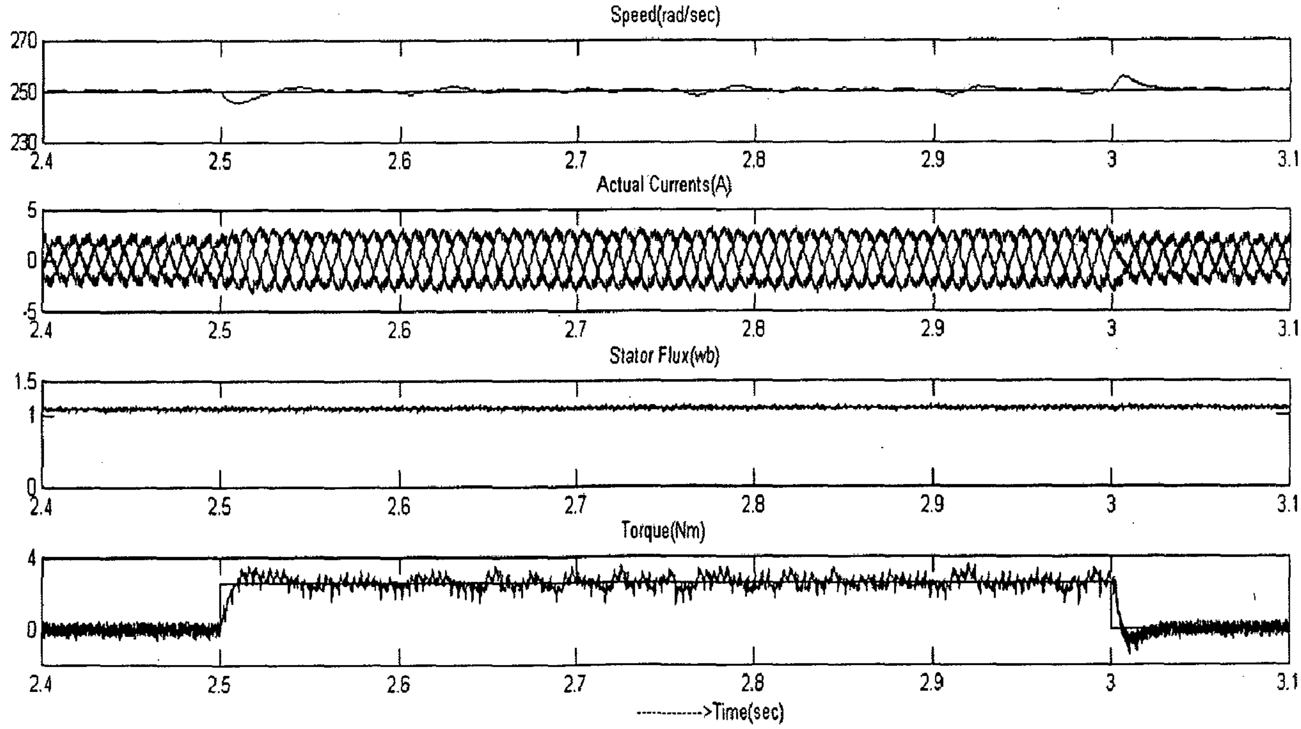

Figure 5.14 Load Perturbation Dynamics of 1HP Direct Torque Controlled Induction Motor Drive System

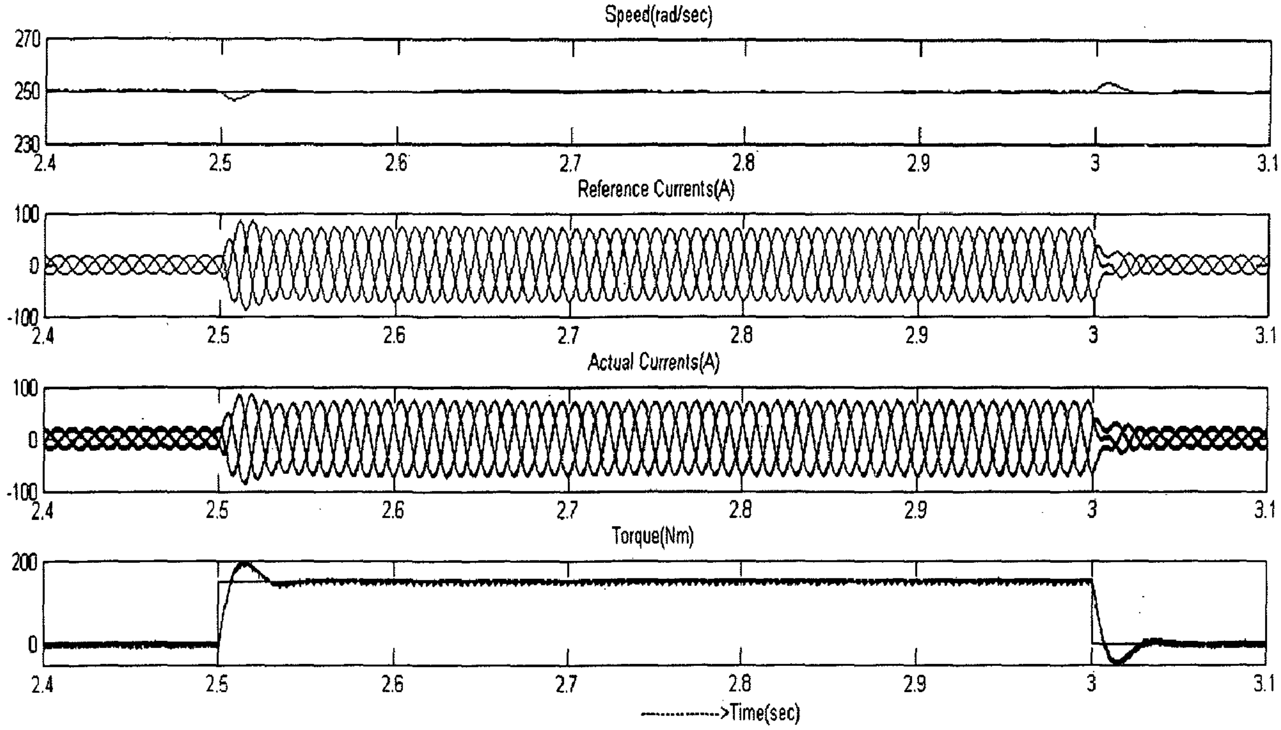

Figure 5.15 Load Perturbation Dynamics of 30HP Vector Controlled Induction Motor Drive System

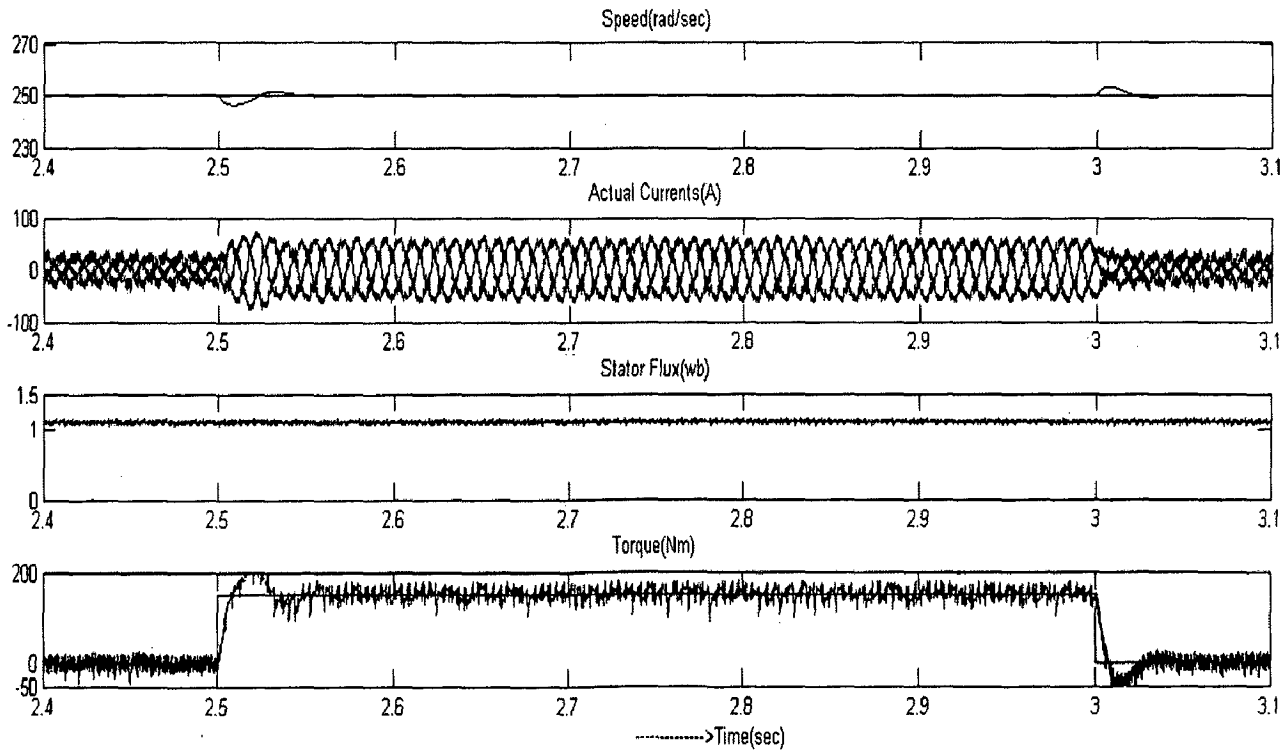

Figure 5.16 Load Perturbation Dynamics of 30HP Direct Torque Controlled Induction Motor Drive System

The comparison results of both the control techniques for 1 HP and 30 HP Induction Motor

Technique	Starting Time (msec)	Reversal Time (msec)	Speed Dip on Full Load Application (rad/sec)	Speed Rise on Full Load Removal (rad/sec)
Vector Control	122	202	3	3
Direct Torque Control	118	199	5	5

Table 5.1 Comparative Performance of Vector Control and Direct Torque Control for 1HP with reference speed of 250 rad/sec, reversal speed of -250 rad/sec and load torque of 2.5Nm

Technique	Starting Time (msec)	Reversal Time (msec)	Speed Dip on Full Load Application (rad/sec)	Speed Rise on Full Load Removal (rad/sec)
Vector Control	190	315	3.1	3.1
Direct Torque Control	182	305	3.6	3.4

Table 5.2 Comparative Performance of Vector Control and Direct Torque Control for 30HP with reference speed of 250 rad/sec, reversal speed of -250 rad/sec and load torque of 150Nm

5.4 COMPARATIVE STUDY

The performance of 0.75KW and 22KW vector control induction motor drive(VCIMD) and direct torque controlled induction motor drive (DTCIMD) system have been simulated under various dynamic conditions namely starting, speed reversal and load perturbation (load application and load removal) in MATLAB environment using simulink and powersystem toolbox tools respectively. The obtained results are shown in Figures. 5.1 to 5.16 and the observed data is tabulated in Tables 5.1 and 5.2. The following observations are made from the results.

A comparative study is carried out between the two control techniques mainly emphasizing on the dynamic performance of the drive. Also we can examine the conceptual complexity of the two schemes in a generalized manner.

In case of the indirect vector control technique considered here, the algorithm involves the decoupled components of the stator current space vector in the synchronously rotating reference frame (SRRF being aligned with the rotor flux) fro the control of flux and torque respectively in a decoupled manner. It also computes the flux angle so that reference decoupled current components can be transferred to the stationary reference frame (SRF). Error between actual and reference currents leads to the generation of PWM signals to operate the VSI in current controlled mode.

In case of the direct torque controlled induction motor drive, the computation complexity simplifies to a great extent. In this control scheme, the control is straight away implemented on the stator flux space vector by optimal selection of the voltage space vector to control VSI. The closed loop control is dependent upon the error between the reference and estimated values of torque and flux respectively. Difficulty in control of torque and flux at low speeds, lack of direct current control are some of the problems that are observed in DTC mode of operation of induction motor drives.

According to the data tabulated from the comparison of the dynamic performance of the induction motor drive which is considered as a very important criteria in the basic control methodology, it is observed that the time required by the induction motor drive to start from rest to a set No Load reference speed and the time required to change from a set reference speed at steady state to the same speed on opposite direction, the method of DTC consumes lesser amount of time as compared to the vector control technique.

Similarly, the speed dip occurring on sudden application of load and overshoot occurring on instantaneous removal of the load is more in case of direct torque controlled induction motor drive. The above observations are checked for two different ratings of induction motors namely 1HP (0.75KW) and 30HP (22KW) respectively.

Henceforth, depending upon various criteria such as basic control, system dynamics and implementation complexity both the techniques are comparable to each other.

5.5 CONCLUSION

After a detailed study of the control methodology of vector control induction motor drive (VCIMD) and direct torque controlled induction motor drive (DTCIMD), it is observed that though the computation complexity reduces in the control strategy of DTCIMD, both the techniques are comparable to each other.

CHAPTER VI

CONCLUSIONS AND FUTURE SCOPE

A comparative study of Vector Controlled Induction Motor Drive (VCIMD) and Direct Torque Controlled Induction Motor Drive (DTCIMD) has been carried out on the basis of control strategy, dynamic performance and probable implementation complexity. The induction motor drive is operated in various modes of operation such as starting, reversal and load perturbation (load application and load removal). The simulation observations stress upon the system dynamics and express the comparative response in terms of starting time, reversal time, speed dip on load application and speed rise on load removal.

The main emphasis has been put on comparison between the two techniques on the basis of system dynamics. It was observed that induction motor drive took lesser time for starting and speed reversal when operated in direct torque control mode as compared to when operated in vector control mode. The speed dip on load application and speed rise on load removal are comparatively lesser in vector control mode of operation.

There are certain factors to be taken into consideration in order to develop a good control technique. The reduction of sensors in both the control techniques can further be investigated without any loss in quality of the estimated signal and therefore the controller can become more robust in nature. This can reduce the overall cost of the drive system. The control logic used for the speed controller can henceforth be upgraded to a better control logic, such as proportional integral derivative (PID) control, sliding mode control, fuzzy logic control etc. The direct torque control technique can also be further improved by the reduction of the torque and current ripples by using an improved optimal switching table.

Henceforth both the control techniques namely field oriented control (FOC) and direct torque control (DTC) schemes have a comparable performance and the choice of the control technique would depend mainly on the requirements of the application.

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APPENDIX – A

MOTOR PARAMETERS-I

Type : Squirrel cage induction motor

Phase : 3

Power : 1 Hp

Voltage : 420 V

Current : 2 A

Speed : 2820 RPM

Poles : 2

Frequency: 50 Hz

Equivalent circuit parameters:

DC Resistance	= 9.27 Ω
Stator resistance per phase	= 11.124 Ω
Rotor resistance per phase	= 8.9838 Ω (referred to stator)
Stator reactance per phase	= 10.48 Ω
Rotor reactance per phase	= 10.48 Ω (referred to stator)
Magnetizing reactance per phase	= 154.08 Ω
Moment of inertia per phase	= 0.0018 J (Kg-m^2)/sec

MOTOR PARAMETERS-II

Type : Squirrel cage induction motor

Phase : 3

Power : 30 Hp

Voltage : 415 V

Current : 45 A

Poles : 4

Frequency: 50 Hz

Equivalent circuit parameters:

Stator resistance per phase	= 0.251Ω
Rotor resistance per phase	= 0.249 Ω (referred to stator)
Stator reactance per phase	= 0.439 Ω
Rotor reactance per phase	= 0.439 Ω (referred to stator)
Magnetizing reactance per phase	= 13.085 Ω
Moment of inertia	= 0.305 J (Kg-m^2)/sec

APPENDIX – B

Six Active and Two Non Active Inverter Switching Vectors