PERFORMANCE INVESTIGATION OF FUZZY ADAPTIVE VECTOR CONTROLLED INDUCTION MOTOR DRIVE

A DISSERTATION

Submitted in partial fulfillment of the requirements for the award of the degree
of
MASTER OF TECHNOLOGY
in
ELECTRICAL ENGINEERING
(With Specialization in Power Apparatus & Electric Drives)

By

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CANDIDATE'S DECLARATION

I hereby declare that the work that is being presented in this dissertation report entitled "**PERFORMANCE INVESTIGATION OF FUZZY ADAPTIVE VECTOR CONTROLLED INDUCTION MOTOR DRIVE**" 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,** to the **Department Of Electrical Engineering, Indian Institute Of Technology, Roorkee,** is an authentic record of my own work carried out, under the guidance of **Dr. S.P.Singh,** Professor, Department of Electrical Engineering.

The matter embodied in this dissertation report has not been submitted by me for the award of any other degree or diploma.

Date: 26/06/08

Place: Roorkee

Kolla Jeevan Kumar

(Kolla Jeevan Kumar)

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

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Date: 26th June, 2008 KOLLA JEEVAN KUMAR

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ABSTRACT

The control and estimation of A.C drives in general are considerably more complex than those of dc drives, and this complexity increases substantially if high performances are demanded. The main reasons for this complexity are the need of variable frequency, harmonically optimum converter power supplies, and the complex dynamics of ac machines, machine parameter variations and difficulties in processing feed back signals in the presence of harmonics. With the advancements of power electronics, digital signal processor technology and different control strategies, a.c drives becomes more efficient.

Indirect field-oriented control (IFOC) method is widely used for IM drives. By providing decoupling of torque and flux control demands, the vector control can navigate an AC motor drive similar to a separately excited DC motor drive without sacrificing the quality of the dynamic performance. Within this scheme, a rotational transducer such as a tachogenerator, an encoder or a resolver, was often mounted on the IM shaft. However, a speed sensor cannot be mounted in some cases, such as motor drives in a hostile environment or high-speed motor drives. Also such sensors lower the system reliability and require special attention to noise. Therefore, sensors less induction motor (IM) drives are widely used in industry for their reliability and flexibility, particularly in hostile environment.

In this work, a comprehensive mathematical modeling of vector controlled induction motor drive (VCIMD) system with speed sensor and with out speed sensor has been carried out to investigate the performance of drive system. VCIMD has been implemented using both Fuzzy and PI controllers. The dynamic response of VCIMD under various operating conditions such as starting, speed reversal and load perturbation is simulated and examined in MATLAB 7 environment using Simulink and power system block set toolboxes.

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CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION

Induction motors are relatively rugged and inexpensive machines. Therefore much attention is given to their control for various applications with different control requirements. An induction machine, especially squirrel cage induction machine, has many advantages when compared with DC machine. First of all, it is very cheap. Next, it has very compact structure and insensitive to environment. Furthermore, it does not require periodic maintenance like DC motors. However, because of its highly non-linear and coupled dynamic structure, an induction machine requires more complex control schemes than DC motors. Traditional open-loop control of the induction machine with variable frequency may provide a satisfactory solution under limited conditions. However, when high performance dynamic operation is required, these methods are unsatisfactory. Therefore, more sophisticated control methods are needed to make the performance of the induction motor comparable with DC motors. Recent developments in the area of drive control techniques, fast semiconductor power switches, powerful and cheap microcontrollers made induction motors alternatives of DC motors in industry.

The most popular induction motor drive control method has been the field oriented control (FOC) in the past two decades. Furthermore, the recent trend in FOC is towards the use of sensorless techniques that avoids the use of speed sensor and flux sensor. The sensors in the hardware of the drive are replaced with state observers to minimize the cost and increase the reliability.

Controlled induction motor drives without mechanical speed sensors at the motor shaft are attraction of low cost and high reliability. To replace the sensor, the information on the rotor speed is extracted from measured stator voltages and currents at the motor terminals. Vector controlled drives require estimating the magnitude and spatial Orientation of the fundamental magnetic flux waves in the stator or in the rotor.

1.2 INDUCTION MACHINE CONTROL

The controllers required for induction motor drives can be divided into two major types: conventional low cost volts per hertz v/f controller and torque controller [1]-[5]. In v/f control, the magnitudes of the voltage and frequency are kept in proportion. The performance of the v/f control is not satisfactory, because the rate of change of voltage and frequency has to be low. A sudden acceleration or deceleration of the voltage and frequency can cause a transient change in the current, which can result in drastic problems. Some efforts were made to improve v/f control performance, but none of these improvements could yield a v/f torque controlled drive systems and this made DC motors a prominent choice for variable speed applications.

This began to change when the theory of field orientation was introduced by Hasse and Blaschke [17]. Field orientation control is considerably more complicated than DC motor control. The most popular class of the successful controllers uses the vector control technique because it controls both the amplitude and phase of AC excitation. This technique results in an orthogonal spatial orientation of the electromagnetic field and torque, commonly known as Field Oriented Control (FOC)

1.3 FIELD ORIENTED CONTROL OF INDUCTION MACHINE

The concept of field orientation control is used to accomplish a decoupled control of flux and torque. This concept is copied from dc machine direct torque control that has these requirements [5]:

- An independent control of armature current to overcome the effects of armature winding resistance, leakage inductance and induced voltage;
- An independent control of constant value of flux;

If all of these requirements are met at every instant of time, the torque will follow the current, allowing an immediate torque control and decoupled flux and torque regulation.

A two phase d-q model of an induction machine rotating at the synchronous speed is introduced which will help to carry out the decoupled control concept to the induction machine and it is quite significant to synthesize the concept of field-oriented control. When three-phase voltages are applied to the machine, they produce three-phase fluxes both in the stator and the rotor. The three-phase fluxes can be represented in a two-phase stationary (α-β) frame. If these two phase fluxes along (α-β) axes are represented by a single-vector then all the machine flux will be aligned along that vector. This vector is commonly specified as d-axis which makes an angle with the stationary frames α-axis, as shown in Fig 1.1. The q-axis is set perpendicular to the d-axis. The flux along the q-axis in this case will be obviously zero. The phasor diagram Fig.1.1 presents these axes.

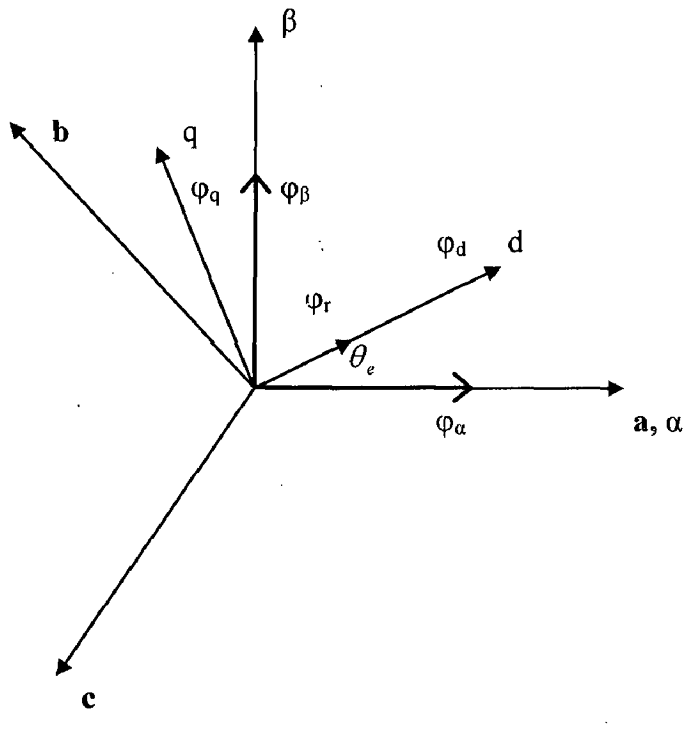

FIGURE.1.1 Phasor Diagram of the Field Oriented Drive System

When the machine input currents change sinusoidally in time, the angle θ_e keeps changing. Thus it is needed to know the angle θ_e accurately, so that the d-axis of the d-q frame is locked with the flux vector. The control inputs can be specified in two phase synchronously rotating d-q frame as i^e_{ds} and i^e_{qs} such that i^e_{ds} being aligned with the d-axis or the flux vector. These two-phase synchronous control inputs are converted into two-an encoder or can be estimated. In case the rotor speed is estimated, the control technique is known as ***sensorless control.***

1.3.1 Indirect Field Oriented Control

In indirect field orientation, the synchronous speed ω_e is the same as the instantaneous speed of the rotor flux vector ψ_{dr} and the d-axis of the d-q coordinate system is exactly locked on the rotor flux vector (rotor flux vector orientation). This facilities the flux control through the magnetizing current i_{ds} by aligning all the flux with the d-axis while aligning the torque-producing component of the current with the q-axis. After decoupling the rotor flux and torque-producing component of the current components, the torque can be instantaneously controlled by controlling the current i_{qs} . The requirement to align the rotor flux with the d-axis of the d-q coordinate system means that the flux along the q-axis must be zero.

Based on this restriction ω_{sl} is:

$$\omega_{sl} = \frac{\frac{1}{T_r} i_{qs}}{\frac{1}{1 - pT_r} i_{ds}} \qquad \text{[ref.no.2]} \qquad (1.1)$$

These relations suggest that flux and torque can be controlled independently by specifying d-q axis currents provided the slip frequency is satisfied (1.1) at all instants.

1.3.2 Direct Field Oriented Control

The DFO control and sensorless control rely heavily on accurate flux estimation. DFOC is most often used for sensorless control, because the flux observer used to estimate the synchronous speed or angle can also be used to estimate the machine speed. Investigation of ways to estimate the flux and speed of the induction machine has also been extensively studied in the past two decades. Classically, the rotor flux was measured by using a special sensing element, such as Hall effect sensors placed in the air-gap. An advantage of this method is that additional required parameters, L_{lr},L_m, and L_r are not

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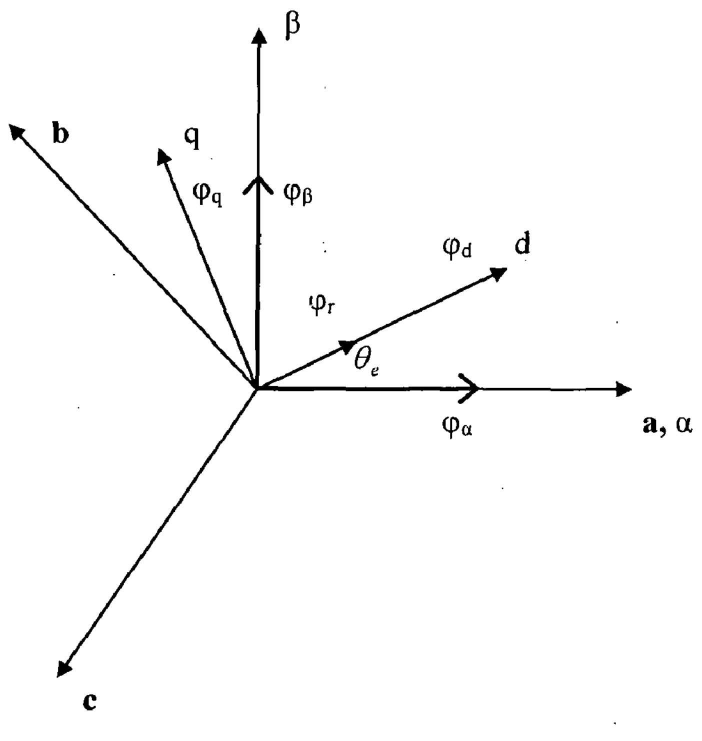

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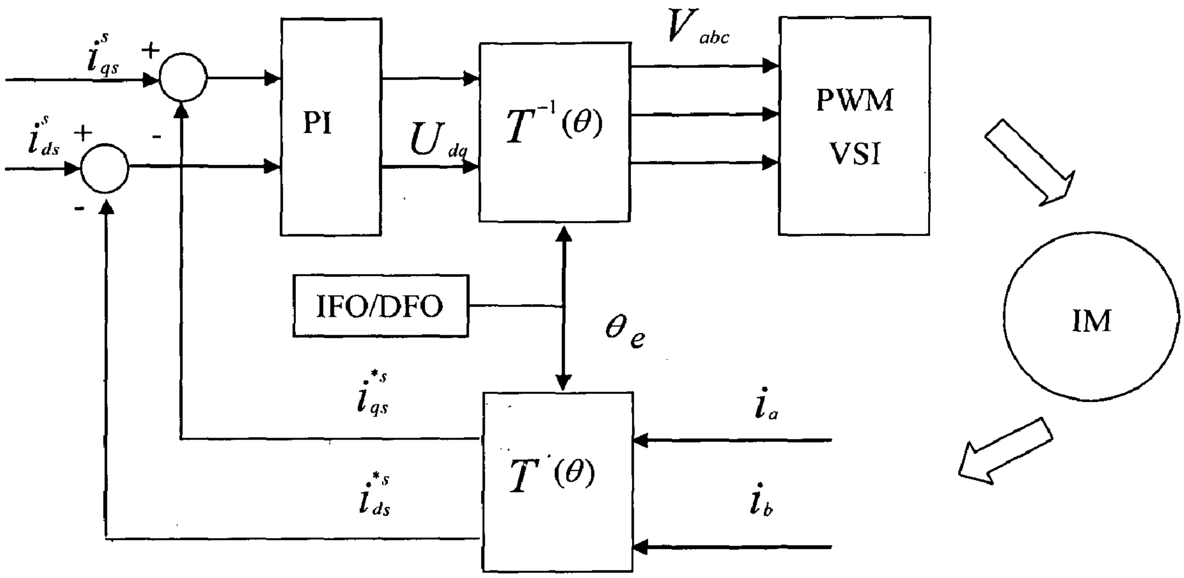

FIGURE.1.2 Field Oriented induction Motor Drive

In the field-oriented controller the flux can be regulated in the stator, air-gap or rotor flux orientation [1]-[5].

The control algorithm for calculation of the rotor flux angle using IFO is based on the assumption that, the flux along the q-axis is zero, which forces the command slip velocity to be $\omega_{sl} = i^s_{qs} /(T_r . i^s_{ds})$ as a necessary and sufficient condition to guarantee that all the flux is aligned with d-axis and the flux along q-axis is zero. The angle θ_e can then be determined as the sum of the slip and the rotor angles after integrating the respective velocities. This slip angle includes the necessary and sufficient condition for decoupled control of flux and torque. The rotor speed can be measured directly by using an encoder or can be estimated. In case the rotor speed is estimated, the control technique is known as ***sensorless control.***

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1.3.3 Sensor less Vector control

In order to implement the vector control technique, the motor speed information is required. Tacho generators, resolvers or incremental encoders are used to detect the rotor speed. However, these sensors impair the ruggedness, reliability and simplicity of the IM. Moreover, they require careful mounting and alignment and special attention is required with electrical noises. Speed sensor needs additional space for mounting and maintenance and hence increases the cost and the size of the drive system. However, in one aspect, the speed sensor elimination reduces the total cost of the drive system. On the other hand the sensorless drive system is more versatile due to the absence of the numerous problems associated with the speed sensor as discussed previously. Therefore it is encouraged to use the sensorless system where the speed is estimated by means of a control algorithm instead of measuring. However eliminating the speed sensor without degrading the performance is still a challenge for engineers.

1.4 LITERATURE REVIEW

The concept of indirect field oriented control developed in the past has been widely studied by researchers during the last two decades. Joseph Vithayathil [11] presented a paper on field oriented control (vector control) of 3 phase induction motor which explains the basic concepts of vector control, as applied to a 3 phase squirrel cage induction motor and also explained in detail and in exact manner, the variables used in the related mathematical theory. The concept of space vectors, which perhaps the simplest means of presenting the concepts of vector control, is also explained in detail.

The rotor flux orientation is both the original and usual choice for the indirect orientation control. Also the IFO control can be implemented in the stator and air-gap flux orientation as well. De Doncker [6] introduced this concept in his universal field oriented controller. In the air-gap flux the slip and flux relations are coupled equations and the d-axis current does not independently control the flux as it does in the rotor flux orientation. For the constant air-gap flux orientation, the maximum torque produced is 20% less than that of the other two methods [4]. In the stator flux orientation, the transient reactance is a coupling factor and it varies with the operating conditions of the machine. In addition, Nasar [4] shows that among these methods, rotor flux oriented control has linear torque curve. Therefore, the most commonly used choice for IFO is the rotor flux orientation.

The IFOC is an open loop, feed-forward control in which the slip frequency is fed-forward guaranteeing the field orientation. This feed-forward control is very sensitive to the rotor open circuit time constant, τ_r. Therefore, τ_r must be known in order to achieve a decoupled control of torque and flux components by controlling, respectively. When τ_r is not set correctly, the machine is said to be detuned and the performance will become sluggish due to loss of decoupled control of torque and flux. The measurement of the rotor time constant, its effects on the system performance and its adaptive tuning to the variations resulting during the operation of the machine have been studied in the literature [6-9]. Lorenz, Krishnan and Novotny [6-9] studied the effect of temperature and saturation level on the rotor time constant and concluded that it can reduce the torque capability of the machine and torque/amps of the machine. The detuning effect becomes more severe in the field-weakening region. Also, it results in a steady-state error and, transient oscillations in the rotor flux and torque. Some of the advanced control techniques such as estimation theory tools and adaptive control tools are also studied to estimate rotor time constant and other motor parameters [20-23].

Speed estimation has greatly evolved from an open loop, low performance strategy to closed loop, high performance strategy over the past decades. The need of developing such technique is essential to adapt to the advancement in the control strategy, especially the vector control techniques. Looking back into the past, Abbondanti [26] has become the first to propose calculating of rotor speed based on the motor model. The fact is that, the real time calculating of the speed has difficulties for the realization because it is largely dependent on the motor's parameters.

Although there are various techniques available for speed sensorless estimation, but not enough effort has been put to review the schemes comparatively. Illas et al. [28] have investigated and compared several speed sensorless estimation schemes for field oriented control of IM drives. Speed estimations using speed estimator, MRAS, speed observer, Kalman filter and rotor slot ripple have been review and simulated to evaluate the performance based on some figures of merit. The studies focus on the level of the difficulty in tuning the adaptive gains and the speed tracking performances. Bodson and Chiasson [29] have considered three representative approaches such as the adaptive method, least-square method and nonlinear method for speed estimation. The methods are compared in terms of their sensitivity to parameters variation, their ability to handle load and their speed tracking capability.

Gabriel [11] avoided the special flux sensors and coils by estimating the rotor flux from the terminal quantities (stator voltages and currents). This technique requires the knowledge of the stator resistance along with the stator, rotor leakage inductances and magnetizing inductance. This method is commonly known as the Voltage Model Flux Observer (VMFO). In this model, integration of the low frequency signals, dominance of stator resistance voltage drop at low speed and leakage inductance variation result in a less precise flux estimation. Integration at low frequency is studied by [12] and three different alternatives are given. Estimation of rotor flux from the terminal quantities depends on parameters such as stator resistance and leakage inductance. The study of parameter sensitivity shows that the leakage inductance can significantly affect the system performance such as stability, dynamic response, and utilizations of the machine and the inverter.

Some studies related to parameter variation effects in sensorless vector controlled drives are already available [36]. For example, impact of rotor resistance variation on transient behavior of the drive was studied by Ilas et al. [28] and by Griva et al. [21] through simulation. Tamai and Schauder [19][26] and Ramukrishan et al. [36] have studied impact of rotor resistance, stator resistance and mutual inductance variation in low speed region experimentally. The only available comprehensive investigations of

steady-state speed estimation errors caused by parameter variation effects appear to be works by Gimenez et al. [20] and Jansen and Lorenz [23]. However, in both cases structure of the drive dealt with is direct rotor flux oriented control that combines a MRAS based speed estimator with a closed loop flux observer and includes a mechanical system model. The validity of results obtained by Gimenez and Jansen is thus restricted to that specific drive structure.

Artificial Intelligence (AI) techniques, such as expert system (ES), fuzzy logic (FL), artificial neural network (ANN), and genetic algorithm (GA) have recently expanded the frontier of power electronics and motor drives which are already complex and interdisciplinary areas. Fuzzy logic and neural network constitute the most important elements of AI and are often defined as "soft computing" or approximate computation compared to hard or precise computation. According to George Boole, who published the article "Investigation of the Laws of Thoughts" in 1854, human beings think and take decisions on the basis of logical "yes/no" or "true/false" reasoning. This gave birth to Boolean algebra and gradually became the basis of our modern digital computers. Lofti Zadeh [28] argued that human thinking is often fuzzy or uncertain in nature and does not always follow crisp "yes/no" logic. He organized the "fuzzy logic" theory in 1965 and created a storm of controversy in the intellectual community. Now fuzzy logic is often applied in the robust control of a feedback system with parameter variation problem and the load torque disturbance.

Various software packages for simulation have been proposed in the recent years, for electronic circuits, like SPICE and SABER, power networks like EMTP and EUROSTAG or specialized simulation of power electronic systems like SIMPLORER, POSTMAC, SIMSEN, ANSIM and PSCAD. Such software packages offer a more or less user friendly environment, but not using the concept of visual design. More recently, a lot of attention have been given to libraries of models for the various components of a power electronic system and electrical machines, developed in the MATLAB/Simulink environment, in order to explore the computational power and flexibility and integrate the visual design facilities.

1.5 ORGANIZATION OF THE DISSERTATION

The main aim of the dissertation is to develop a control strategy for an induction motor so as to get the desired performance without using the speed sensor (i.e. sensor less speed control of induction motor).This dissertation is organized as follows:

The first chapter introduces the problem envisaged based on the literature review presented in the chapter in the field of vector control of induction motor.

The second chapter deals with a generalized dynamic mathematical model of the induction motor which can be used to construct various equivalent circuit models in different reference frames.

The third chapter presents simulink implementation of the induction motor in stationary frame and indirect vector control and the Fuzzy Logic control are discussed.

The fourth chapter is dealt with Full-order speed adaptive observer and vector control for sensor less control of Induction motor drive.

The fifth chapter includes the results and discussions of the Vector control with sensor and without sensor using Fuzzy and PI controllers.

The sixth chapter concludes the work done and suggestions for further work in the field of induction motor control technology are presented.

In the last, the appendices are given, which form a part of the chapters discussed in the report.

DYNAMIC MODEL OF THREE-PHASE INDUCTION MOTOR

2.1 INTRODUCTION

The two names for the same type of motor, *Induction motor* and *Asynchronous motor*, describe the two characteristics in which this type of motor differs from DC motors and synchronous motors. Induction refers to the fact that the field in the rotor is induced by the stator currents, and asynchronous refers to the fact that the rotor speed is not equal to the stator frequency. No sliding contacts and permanent magnets are needed to make an induction motor work, which makes it very simple and cheap to manufacture. As motors, they rugged and require very little maintenance. However, their speeds are not as easily controlled as with DC motors. They draw large starting currents, and operate with a poor lagging factor when lightly loaded.

The present chapter deals with the dynamic model of a 3-Φ Induction motor particularly of squirrel cage type.

2.2 DYNAMIC MODEL OF A THREE-PHASE INDUCTION MOTOR

Most induction motors are of the rotary type with basically a stationary stator and a rotating rotor. The stator has a cylindrical magnetic core that is housed inside a metal frame. The stator magnetic core is formed by stacking thin electrical steel laminations with uniformly spaced slots stamped in the inner circumference to accommodate the three distributed stator windings. The stator windings are formed by connecting coils of copper or aluminum conductors that are insulated from the slot walls.

The rotor consists of a cylindrical laminated iron core with uniformly spaced peripheral slots to accommodate the rotor windings. In this thesis a squirrel cage rotor induction motor is used. It has uniformly spaced axial bars that are soldered onto end rings at both ends. After the rotor core laminations are stacked in a mold, the mold is filled with molten aluminum.

During the entire report, a complex vector notation and some reference frame conversions are used.

2.2.1 Mathematical Model Of Induction Motor

A two-pole, three-phase symmetrical induction machine is considered as shown in Fig. 2.1. The stator windings are identical, sinusoidally distributed winding displaced by 120^0, with N_s equivalent turns and resistance r_s. The rotor winding having three identical sinusoidal distributed windings displaced $120^{0,}$ with N_r equivalent turns and resistance r_r.

In this analysis, the implicit assumption is that stator and rotor self inductance of the induction machine can be represented as sum of lumped value of leakage inductance and magnetizing inductance. Leakage inductance is assumed to be constant always.

The idealized three-phase induction motor is assumed to have symmetrical air gap. The two common qd0 reference frames in the analysis of induction machine at the stationary and the synchronously rotating reference frame. In the stationary rotating reference, the dqo variables of the machine are in the same frame as those being used for supply network. It is a convenient choice of frame when the supply network is large or complex .For the transient study, the induction machine is simulated on a stationary reference frame.

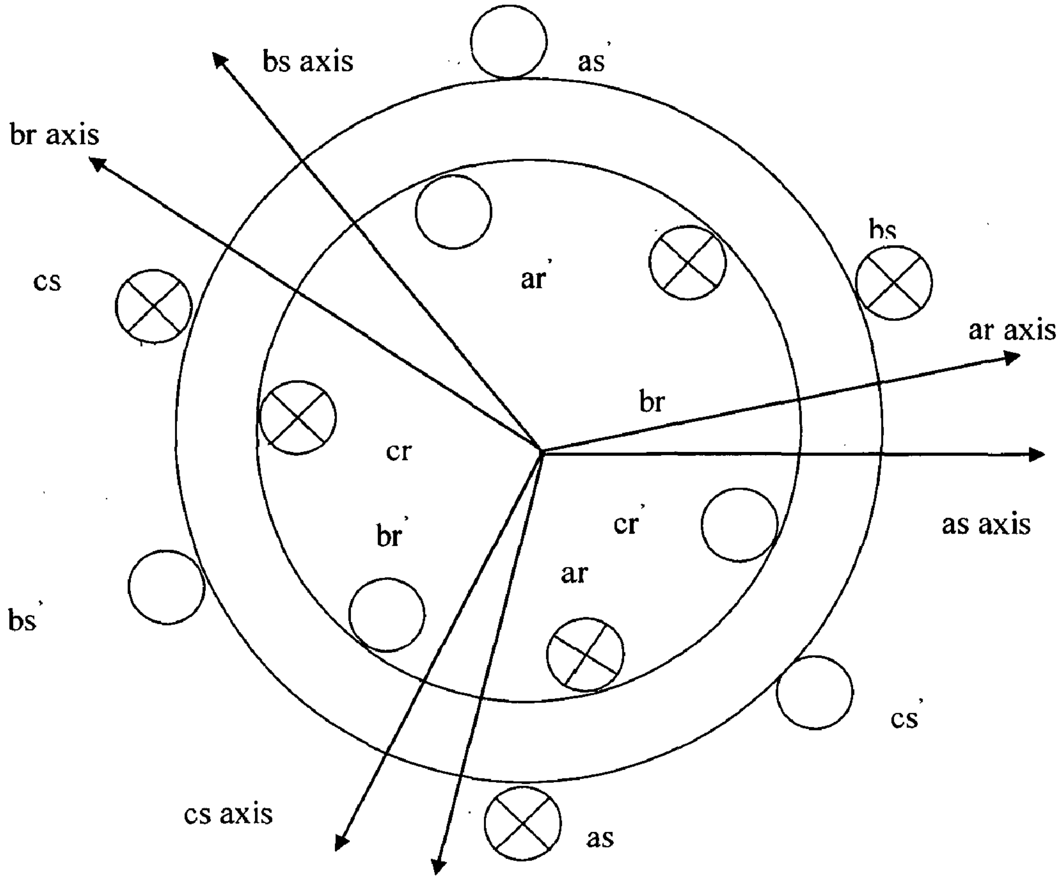

FIGURE.2.1 Two-pole, 3-Φ, symmetrical induction machine

First deriving the equations of the induction machines in the arbitrary reference frame which is rotating at a speed of ω in the direction of rotor rotation and then setting ω=0 , equations of the machine in the stationary reference frame are obtained and setting $\omega = \omega_e$ equations of the machine in the synchronously rotating reference frame is obtained.

2.2.1.1 **Development of Voltage and Torque Equations**

This section deals with the development of voltage and torque equations in arbitrary reference frame for symmetrical three-phase induction machine. Since in the analysis of the induction machine all the time varying inductances must be eliminated, it is necessary to transform all machine variables to stationary reference frame in order to obtain voltage equation with constant parameters.

The induction motors are of a rotary type with basically a stationary stator and a rotating rotor. The voltage equation of the magnetically coupled stator & rotor circuits can be written as follows [1]:

Stator Voltage Equation:

$$v_{as} = r_s i_{as} + p\lambda_{as} \tag{2.1a}$$

$$v_{bs} = r_s i_{bs} + p\lambda_{bs} \tag{2.1b}$$

$$v_{cs} = r_s i_{cs} + p\lambda_{cs} \tag{2.1c}$$

where the subscript's' stands for stator circuits

Rotor Voltage Equation:

$$v_{ar} = r_r i_{ar} + p\lambda_{ar} \tag{2.2a}$$

$$v_{br} = r_r i_{br} + p\lambda_{br} \tag{2.2b}$$

$$v_{cr} = r_r i_{cr} + p\lambda_{cr} \tag{2.2c}$$

where the subscripts 'r' refers to rotor quantities.

Flux Linkage Equations:

The flux linkage of the stator and rotor winding, in terms of the winding inductances and currents is written as:

$$\begin{bmatrix} \lambda abcs \\ \lambda abcr \end{bmatrix} = \begin{bmatrix} Ls & Lsr \\ (Lsr)^T & Lr \end{bmatrix} \begin{bmatrix} i abcs \\ i abcr \end{bmatrix} \tag{2.3}$$

where

$$L_s = \begin{bmatrix} L_{ls} + L_{ms} & -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} \\ -\frac{1}{2}L_{ms} & L_{ls} + L_{ms} & -\frac{1}{2}L_{ms} \\ -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} & L_{ls} + L_{ms} \end{bmatrix} \tag{2.4}$$

$$L_r = \begin{bmatrix} L_{lr} + L_{mr} & -\frac{1}{2}L_{mr} & -\frac{1}{2}L_{mr} \\ -\frac{1}{2}L_{mr} & L_{lr} + L_{mr} & -\frac{1}{2}L_{mr} \\ -\frac{1}{2}L_{mr} & -\frac{1}{2}L_{mr} & L_{lr} + L_{mr} \end{bmatrix} \tag{2.5}$$

$$L_{sr} = L_{sr} \begin{bmatrix} \cos\theta_r & \cos\left(\theta_r + \frac{2\pi}{3}\right) & \cos\left(\theta_r - \frac{2\pi}{3}\right) \\ \cos\left(\theta_r - \frac{2\pi}{3}\right) & \cos\theta_r & \cos\left(\theta_r + \frac{2\pi}{3}\right) \\ \cos\left(\theta_r + \frac{2\pi}{3}\right) & \cos\left(\theta_r - \frac{2\pi}{3}\right) & \cos\theta_r \end{bmatrix} \tag{2.6}$$

where,

L_{ls} = per phase stator winding leakage inductance.

L_{lr} = per phase rotor winding leakage inductance

L_{ms} = magnetizing inductance of the stator winding

L_{mr} = magnetizing inductance of the rotor winding.

L_{sr} = mutual inductance of stator to rotor windings.

When expressing the voltage equations in machine variable it is convenient to refer all rotor variables to the stator windings by appropriate turn's ratios.

$$i'_{abcr} = \frac{N_r}{N_s} i_{abcr} \tag{2.7}$$

$$v'_{abcr} = \frac{N_s}{N_r} v_{abcr} \tag{2.8}$$

$$\lambda'_{abcr} = \frac{N_s}{N_r} \lambda_{abcr} \tag{2.9}$$

The magnetizing and mutual inductances are associated with the same magnetic flux path.

Therefore L_{ms}, L_{mr} and L_{sr} are related as

$$L_{ms} = \frac{N_s}{N_r} L_{sr} \tag{2.10}$$

Thus we will define

$$L'_{sr} = \frac{N_s}{N_r} L_{sr} \tag{2.11}$$

$$= L_{ms} \begin{bmatrix} \cos\theta_r & \cos\left(\theta_r + \frac{2\pi}{3}\right) & \cos\left(\theta_r - \frac{2\pi}{3}\right) \\ \cos\left(\theta_r - \frac{2\pi}{3}\right) & \cos\theta_r & \cos\left(\theta_r + \frac{2\pi}{3}\right) \\ \cos\left(\theta_r + \frac{2\pi}{3}\right) & \cos\left(\theta_r - \frac{2\pi}{3}\right) & \cos\theta_r \end{bmatrix} \tag{2.12}$$

$$L_{mr} = \left(\frac{N_r}{N_s}\right)^2 L_{sr}$$

and if we let $L'_r = \left(\frac{N_s}{N_r}\right)^2 L_r$; Rotor inductance referred to the Stator.

then from Eq. (2.5)

$$L'_r = \begin{bmatrix} L'_{lr} + L_{ms} & -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} \\ -\frac{1}{2}L_{ms} & L'_{lr} + L_{ms} & -\frac{1}{2}L_{ms} \\ -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} & L'_{lr} + L_{ms} \end{bmatrix} \quad (2.13)$$

where $L'_{lr} = \left(\frac{N_s}{N_r}\right)^2 L_{lr}$; (2.14)

L'_{lr} is Rotor leakage inductance referred to stator

Rotor inductance referred to the stator

The flux linkages may now be expressed as

$$\begin{bmatrix} \lambda_{abcs} \\ \lambda'_{abcr} \end{bmatrix} = \begin{bmatrix} L_s & L'_{sr} \\ (L'_{sr})^T & L'_r \end{bmatrix} \begin{bmatrix} i_{abcs} \\ i'_{abcr} \end{bmatrix} \quad (2.15)$$

The voltage equations expressed in terms of machine variables referred to the stator windings may now be written as

$$\begin{bmatrix} V_{abcs} \\ V'_{abcr} \end{bmatrix} = \begin{bmatrix} r_s + pL_s & pL'_{sr} \\ p(L'_{sr})^T & r'_r + pL'_r \end{bmatrix} \begin{bmatrix} i_{abcs} \\ i'_{abcr} \end{bmatrix} \quad (2.16)$$

However, electromagnetic torque T_e and speed are related by the following equation

$$T_e = J(\frac{2}{P})p\omega_r + T_L \quad (2.17)$$

where J is the moment of inertia of the rotor and T_L is the connected load torque

The idealized three-phase induction motor is assumed to have symmetrical air gap. The two common qd0 reference frames in the analysis of induction machine at the stationary and the synchronously rotating reference frame. In the stationary rotating reference, the dq variables of the machine are in the same frame as those being used for supply network. It is a convenient choice of frame when the supply network is large or complex .For the transient study, the induction machine is simulated on a stationary reference frame. In the synchronously rotating reference frame the dq variable are in steady state, a prerequisite when deriving a small signal model about a chosen operating point.

First deriving the equations of the induction machines in the arbitrary reference frame which is rotating at a speed of ω in the direction of rotor rotation and then setting ω=0 , equations of the machine in the stationary reference frame are obtained.

The voltage equations in the arbitrary reference frame are

$$V_{qdos} = r_s i_{qdos} + \omega\lambda_{dqs} + p\lambda_{dqs} \tag{2.18}$$

$$V'_{qdor} = r'_r i'_{qdor} + (\omega - \omega_r)\lambda'_{dqr} + p\lambda'_{qdor} \tag{2.19}$$

where

$$\begin{aligned}(\lambda_{dqs})^T &= \begin{bmatrix}\lambda_{ds} & -\lambda_{qs} & 0\end{bmatrix}\\ (\lambda'_{dqr})^T &= \begin{bmatrix}\lambda'_{dr} & -\lambda'_{qr} & 0\end{bmatrix}\end{aligned} \tag{2.20}$$

The set of equations is complete once the expressions for the flux linkages are determined.

The above two equations can also be written as

$$(\lambda_{qd0s}) = \begin{bmatrix} L_{ls}+M & 0 & 0 \\ 0 & L_{ls}+M & 0 \\ 0 & 0 & L_{ls} \end{bmatrix}\begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{0s} \end{bmatrix} + \begin{bmatrix} M & 0 & 0 \\ 0 & M & 0 \\ 0 & 0 & 0 \end{bmatrix}\begin{bmatrix} i'_{qr} \\ i'_{dr} \\ i'_{0r} \end{bmatrix} \tag{2.21}$$

$$(\lambda'_{qdor}) = \begin{bmatrix} M & 0 & 0 \\ 0 & M & 0 \\ 0 & 0 & 0 \end{bmatrix}\begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{0s} \end{bmatrix} + \begin{bmatrix} L'_{lr}+M & 0 & 0 \\ 0 & L'_{lr}+M & 0 \\ 0 & 0 & L'_{lr} \end{bmatrix}\begin{bmatrix} i'_{qr} \\ i'_{dr} \\ i'_{0r} \end{bmatrix} \tag{2.22}$$

where M=3/2 L_{ms}.

where the prime (') denotes rotor quantities referred to stator side.

The voltage equations are often written in expanded form from (2.18) & (2.19)

$$
\left.\begin{aligned}
v_{qs} &= p\lambda_{qs} + \omega\lambda_{ds} + r_s i_{qs} \\
v_{ds} &= p\lambda_{qs} - \omega\lambda_{qs} + r_s i_{ds} \\
v_{0s} &= p\lambda_{0s} + r_s i_{0s} \\
v'_{qr} &= r'_r i'_{qr} + (\omega - \omega_r)\lambda'_{dr} + p\lambda'_{qr} \\
v'_{dr} &= r'_r i'_{dr} - (\omega - \omega_r)\lambda'_{qr} + p\lambda'_{dr} \\
v'_{0r} &= r'_r i'_{0r} + p\lambda'_{0r}
\end{aligned}\right] \qquad (2.23)
$$

From equation (2.21) & (2.22) flux linkages in expanded form

$$
\left.\begin{aligned}
\lambda_{qs} &= L_{ls} i_{qs} + M(i_{qs} + i'_{qr}) \\
\lambda_{ds} &= L_{ls} i_{ds} + M(i_{ds} + i'_{dr}) \\
\lambda_{0s} &= L_{ls} i_{0s} \\
\lambda'_{qr} &= L'_{lr} i'_{qr} + M(i_{qs} + i'_{qr}) \\
\lambda'_{dr} &= L'_{lr} i'_{qr} + M(i_{ds} + i'_{dr}) \\
\lambda'_{0r} &= L'_{lr} i'_{0r}
\end{aligned}\right] \qquad (2.24)
$$

Since machine and power system parameters are nearly always given in ohms or percent or per unit of base impedances, it is rather convenient to express the voltage and flux linkage equations in terms of reactance rather than inductances. Hence the equations above are written as

$$
\left.
\begin{aligned}
v_{qs} &= r_s i_{qs} + \frac{p}{\omega_b}\psi_{qs} + \frac{\omega}{\omega_b}\Psi_{ds} \\
v_{ds} &= r_s i_{ds} + \frac{p}{\omega_b}\psi_{ds} - \frac{\omega}{\omega_b}\psi_{qs} \\
v_{0s} &= r_s i_{0s} + \frac{p}{\omega_b}\psi_{0s}
\end{aligned}
\right] \tag{2.25a}
$$

$$
\left.
\begin{aligned}
v'_{qr} &= r'_r i'_{qr} + \frac{p}{\omega_b}\psi'_{qr} + \left(\frac{\omega-\omega_r}{\omega_b}\right)\psi'_{dr} \\
v'_{dr} &= r'_r i'_{qr} + \frac{p}{\omega_b}\psi'_{dr} - \left(\frac{\omega-\omega_r}{\omega_b}\right)\psi'_{qr} \\
v'_{0r} &= r'_r i'_{0r} + \frac{p}{\omega_b}\Psi'_{0r}
\end{aligned}
\right] \tag{2.25b}
$$

where ω_b is the base electrical angular velocity used to calculate the inductive reactance. Flux linkages per sec with units of volts are rewritten as

$$
\left.
\begin{aligned}
\psi_{qs} &= x_{ls} i_{qs} + X_M(i_{qs} + i'_{qr}) \\
\psi_{ds} &= x_{ls} i_{ds} + X_M(i_{ds} + i'_{dr}) \\
\psi_{0s} &= x_{ls} i_{0s} \\
\psi'_{qr} &= x'_{lr} i'_{qr} + X_M(i_{qs} + i'_{qr}) \\
\psi'_{dr} &= x'_{lr} i'_{dr} + X_M(i_{ds} + i'_{dr}) \\
\psi'_{0r} &= x'_{lr} i'_{0r}
\end{aligned}
\right] \tag{2.26}
$$

The equations (2.23) and (2.25) are written in terms of currents and flux linkages (flux linkages per sec.).Clearly the currents &flux linkages are related and both can't be independent or state variables. In transfer function formulation or computer simulation of induction motor we will find it is desirable to express voltage equations in terms of either currents or flux linkages (flux linkages per sec.).

If the currents are selected as independent variables and flux linkages or flux linkages per sec. replaced by the currents, the voltage equations become

$$\begin{bmatrix} v_{qs} \\ v_{ds} \\ v_{0s} \\ v'_{qr} \\ v'_{dr} \\ v'_{0r} \end{bmatrix} = \begin{bmatrix} r_s + \frac{p}{\omega_b} X_{ss} & \frac{\omega}{\omega_b} X_{ss} & 0 & \frac{p}{\omega_b} X_M & \frac{\omega}{\omega_b} X_M & 0 \\ -\frac{\omega}{\omega_b} X_{ss} & r_s + \frac{p}{\omega_b} X_{ss} & 0 & -\frac{\omega}{\omega_b} X_M & \frac{p}{\omega_b} X_M & 0 \\ 0 & 0 & r_s + \frac{p}{\omega_b} X_{ls} & 0 & 0 & 0 \\ \frac{p}{\omega_b} X_M & (\frac{\omega - \omega_r}{\omega_b}) X_M & 0 & r'_r + \frac{p}{\omega_b} X'_{rr} & (\frac{\omega - \omega_r}{\omega_b}) X'_{rr} & 0 \\ \left(\frac{\omega_r - \omega}{\omega_b}\right) X_M & \frac{p}{\omega_b} X_M & 0 & \left(\frac{\omega_r - \omega}{\omega_b}\right) X'_{rr} & r'_r + \frac{p}{\omega_b} X'_{rr} & 0 \\ 0 & 0 & 0 & 0 & 0 & r'_r + \frac{p}{\omega_b} X'_{lr} \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{0s} \\ i'_{qr} \\ i'_{dr} \\ i'_{0r} \end{bmatrix}$$

(2.27)

where

$$X_{SS} = X_{ls} + X_M \tag{2.28}$$

$$X'_{rr} = X'_{lr} + X_M \tag{2.29}$$

The Flux linkage per sec. may be expressed from Eq (2.26) as

$$\begin{bmatrix} \Psi_{qs} \\ \Psi_{ds} \\ \Psi_{0s} \\ \Psi'_{qr} \\ \Psi'_{dr} \\ \Psi'_{0r} \end{bmatrix} = \begin{bmatrix} \Psi_{ss} & 0 & 0 & X_M & 0 & 0 \\ 0 & X_{ss} & 0 & 0 & X_M & 0 \\ 0 & 0 & X_{ls} & 0 & 0 & 0 \\ X_M & 0 & 0 & X'_{rr} & 0 & 0 \\ 0 & X_M & 0 & 0 & X'_{rr} & 0 \\ 0 & 0 & 0 & 0 & 0 & X'_{lr} \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{0s} \\ i'_{qr} \\ i'_{dr} \\ i'_{0r} \end{bmatrix} \tag{2.30}$$

If Flux linkage or flux linkages per sec. are selected as independent variables then equation (2.30) may be solved for currents and written as

$$\begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{0s} \\ i'_{qr} \\ i'_{dr} \\ i'_{0r} \end{bmatrix} = \frac{1}{D} \begin{bmatrix} X'_{rr} & 0 & 0 & -X_M & 0 & 0 \\ 0 & X'_{rr} & 0 & 0 & -X_M & 0 \\ 0 & 0 & \dfrac{D}{X_{ls}} & 0 & 0 & 0 \\ -X_M & 0 & 0 & X_{ss} & 0 & 0 \\ 0 & -X_M & 0 & 0 & X_{ss} & 0 \\ 0 & 0 & 0 & 0 & 0 & \dfrac{D}{X'_{lr}} \end{bmatrix} \begin{bmatrix} \Psi_{qs} \\ \Psi_{ds} \\ \Psi_{0s} \\ \Psi'_{qr} \\ \Psi'_{dr} \\ \Psi'_{0r} \end{bmatrix} \qquad (2.31)$$

where $D = X_{ss} X'_{rr} - X_M^2$ (2.32)

Substituting equation (2.31) for currents into (2.25) yields the voltage equations in terms of flux linkages per sec. as

$$\begin{bmatrix} v_{qs} \\ v_{ds} \\ v_{0s} \\ v'_{qr} \\ v'_{dr} \\ v'_{0r} \end{bmatrix} = \begin{bmatrix} \dfrac{r_s X'_{rr}}{D} + \dfrac{p}{\omega_b} & \dfrac{\omega}{\omega_b} & 0 & -\dfrac{r_s X_M}{D} & 0 & 0 \\ -\dfrac{\omega}{\omega_b} & \dfrac{r_s X'_{rr}}{D} + \dfrac{p}{\omega_b} & 0 & 0 & -\dfrac{r_s}{D} X_M & 0 \\ 0 & 0 & \dfrac{r_s}{X_{ls}} + \dfrac{p}{\omega_b} & 0 & 0 & 0 \\ -\dfrac{r'_r X_M}{D} & 0 & 0 & \dfrac{r'_r X_{ss}}{D} + \dfrac{p}{\omega_b} & \left(\dfrac{\omega - \omega_r}{\omega_b}\right) & 0 \\ 0 & -\dfrac{r'_r X_M}{D} & 0 & \left(\dfrac{\omega_r - \omega}{\omega_b}\right) & \dfrac{r'_r X_{ss}}{D} + \dfrac{p}{\omega_b} & 0 \\ 0 & 0 & 0 & 0 & 0 & \dfrac{r'_r}{X'_{lr}} + \dfrac{p}{\omega_b} \end{bmatrix} \begin{bmatrix} \psi_{qs} \\ \psi_{ds} \\ \psi_{0s} \\ \psi'_{qr} \\ \psi'_{dr} \\ \psi'_{0r} \end{bmatrix}$$

(2.33)

It is interesting to note that each q- and d- voltage equation contains two derivatives of current when currents are selected as independent variables or state variables, Eq. (2.27).When flux linkages are selected as independent variables ,(2.33) each q- and d- voltage equation contains only one derivative of flux linkage .In the next chapter we will see that this property makes it more convenient to implement a computer simulation of an induction motor with flux linkages as state variables rather than with currents.

2.3 STATE-SPACE MODEL OF INDUCTION MOTOR

The dynamic model in state-space form is important for transient analysis, particularly for computer simulation study. Although the rotating frame model is generally preferred, stationary frame model can be used. The electrical variables in the model chosen are currents and fluxes. In this part, state-space equations of the machine in rotating frame with flux linkages as main variables.

The flux linkage variables can be defined as

$$\lambda_{qs} = \omega_b \psi_{qs} \tag{2.34}$$

$$\lambda_{qr} = \omega_b \psi_{qr} \tag{2.35}$$

$$\lambda_{ds} = \omega_b \psi_{ds} \tag{2.36}$$

$$\lambda_{dr} = \omega_b \psi_{dr} \tag{2.37}$$

where $\omega_b =$ base frequency of the machine. Substituting these equations in (2.25a)-(2.25b) and for the synchronously rotating frame taking $\omega = \omega_e$, we get

$$v_{qs} = r_s i_{qs} + \frac{1}{\omega_b}\frac{d\lambda_{qs}}{dt} + \frac{\omega_e}{\omega_b}\lambda_{ds} \tag{2.38}$$

$$v_{ds} = r_s i_{ds} + \frac{1}{\omega_b}\frac{d\lambda_{ds}}{dt} - \frac{\omega_e}{\omega_b}\lambda_{qs} \tag{2.39}$$

$$0 = r_r i_{qr} + \frac{1}{\omega_b}\frac{d\lambda_{qr}}{dt} + \frac{(\omega_e - \omega_r)}{\omega_b}\lambda_{dr} \tag{2.40}$$

$$0 = r_r i_{dr} + \frac{1}{\omega_b}\frac{d\lambda_{dr}}{dt} - \frac{(\omega_e - \omega_r)}{\omega_b}\lambda_{qr} \tag{2.41}$$

where it is assumed that $v'_{qr} = v'_{dr} = 0$.

$$\lambda_{qs} = \omega_b \psi_{qs} = X_{ls} i_{qs} + X_m (i_{qs} + i_{qr}) \tag{2.42}$$

$$\lambda_{qr} = \omega_b \psi_{qr} = X_{lr} i_{qr} + X_m (i_{qs} + i_{qr}) \tag{2.43}$$

$$\lambda_{qm} = \omega_b \psi_{qm} = X_m (i_{qs} + i_{qr}) \tag{2.44}$$

$$\lambda_{ds} = \omega_b \psi_{ds} = X_{ls} i_{ds} + X_m (i_{ds} + i_{dr}) \tag{2.45}$$

$$\lambda_{dr} = \omega_b \psi_{dr} = X_{lr} i_{dr} + X_m (i_{ds} + i_{dr}) \tag{2.46}$$

$$\lambda_{dm} = \omega_b \psi_{dm} = X_m (i_{ds} + i_{dr}) \tag{2.47}$$

where $X_{ls} = \omega_b Lls$, $X_{lr} = \omega_b Llr$, and $X_m = \omega_b Lm$, or

$$\lambda_{qs} = X_{ls} i_{qs} + \lambda_{qm} \tag{2.48}$$

$$\lambda_{qr} = X_{lr} i_{qr} + \lambda_{qm} \tag{2.49}$$

$$\lambda_{ds} = X_{ls} i_{ds} + \lambda_{dm} \tag{2.50}$$

$$\lambda_{dr} = X_{lr} i_{dr} + \lambda_{dm} \tag{2.51}$$

From (2.58)-(2.61), the current can be expressed in terms of the flux linkages as

$$i_{qs} = \frac{\lambda_{qs} - \lambda_{qm}}{X_{ls}} \tag{2.52}$$

$$i_{qr} = \frac{\lambda_{qr} - \lambda_{qm}}{X_{lr}} \tag{2.53}$$

$$i_{ds} = \frac{\lambda_{ds} - \lambda_{dm}}{X_{ls}} \tag{2.54}$$

$$i_{dr} = \frac{\lambda_{dr} - \lambda_{dm}}{X_{lr}} \tag{2.55}$$

Substituting (2.52)-(2.55) in (2.48)-(2.49), respectively, we get

$$\lambda_{qm} = \frac{X_{mstar}}{X_{ls}} \lambda_{qs} + \frac{X_{mstar}}{X_{lr}} \lambda_{qr} \tag{2.56}$$

$$\lambda_{dm} = \frac{X_{mstar}}{X_{ls}} \lambda_{ds} + \frac{X_{mstar}}{X_{lr}} \lambda_{dr} \tag{2.57}$$

where

$$X_{mstar} = \frac{1}{\left(\frac{1}{X_m} + \frac{1}{X_{ls}} + \frac{1}{X_{lr}}\right)}$$

Substituting the current equations (2.52)-(2.55) into the voltage equations (2.38)-(2.41),

$$v_{qs} = \frac{r_s}{X_{ls}}\left(\lambda_{qs} - \lambda_{qm}\right) + \frac{1}{\omega_b}\frac{d\lambda_{qs}}{dt} + \frac{\omega_e}{\omega_b}\lambda_{ds} \tag{2.58}$$

$$v_{ds} = \frac{r_s}{X_{ls}}\left(\lambda_{ds} - \lambda_{dm}\right) + \frac{1}{\omega_b}\frac{d\lambda_{ds}}{dt} - \frac{\omega_e}{\omega_b}\lambda_{qs} \tag{2.59}$$

$$0 = \frac{r_r}{X_{lr}}\left(\lambda_{qr} - \lambda_{qm}\right) + \frac{1}{\omega_b}\frac{d\lambda_{qr}}{dt} + \frac{(\omega_e - \omega_r)}{\omega_b}\lambda_{dr} \tag{2.60}$$

$$0 = \frac{r_r}{X_{lr}}\left(\lambda_{dr} - \lambda_{dm}\right) + \frac{1}{\omega_b}\frac{d\lambda_{dr}}{dt} - \frac{(\omega_e - \omega_r)}{\omega_b}\lambda_{qr} \tag{2.61}$$

which can be expressed in state space form as,

$$\frac{d\lambda_{qs}}{dt} = \omega_b\left[v_{qs} - \frac{\omega_e}{\omega_b}\lambda_{ds} - \frac{r_s}{X_{ls}}(\lambda_{qs} - \lambda_{qm})\right] \tag{2.62}$$

$$\frac{d\lambda_{ds}}{dt} = \omega_b\left[v_{ds} - \frac{\omega_e}{\omega_b}\lambda_{qs} - \frac{r_s}{X_{ls}}(\lambda_{ds} - \lambda_{dm})\right] \tag{2.63}$$

$$\frac{d\lambda_{qr}}{dt} = -\omega_b\left[\frac{\left(\omega_e - \omega_r\right)}{\omega_b}\lambda_{dr} + \frac{r_r}{X_{lr}}(\lambda_{qr} - \lambda_{qm})\right] \tag{2.64}$$

$$\frac{d\lambda_{dr}}{dt} = -\omega_b\left[-\frac{\left(\omega_e - \omega_r\right)}{\omega_b}\lambda_{qr} + \frac{r_r}{X_{lr}}(\lambda_{dr} - \lambda_{dm})\right] \tag{2.65}$$

Finally, from equation (2. 36), we get

$$T_e = \left(\frac{3}{2}\right)\left(\frac{P}{2}\right)\left(\frac{1}{\omega_b}\right)\left(\lambda_{ds} i_{qs} - \lambda_{qs} i_{ds}\right) \tag{2.66}$$

$$T_e = T_l + J\frac{d\omega_m}{dt} = T_l + \frac{2}{P}J\frac{d\omega_r}{dt} \tag{2.67}$$

When an electrical machine is simulated in circuit simulators like PSpice, its steady state model is used, but for electrical drive studies, the transient behavior is equally important. One advantage of Simulink over circuit simulators is the ease in modeling the transients of electrical machines and drives and to include drive controls in the simulation.

Simulink induction machine models are available in the literature [14, 18, 19], but that appear to be black-boxes with no internal details. Some of them recommended using S-function, which are software source codes for simulink blocks. This technique does not fully utilize the power and ease of Simulink because S-function programming knowledge is required to access the model variables. S-function runs faster than discrete Simulink blocks, but simulink model can be made to run faster using "accelerator" functions or providing stand-alone Simulink models. Another approach is using the *Simulink Power System Blockset.*

It is easier to develop a basic induction machine and a vector controlled induction machine using MATLAB package, as many components of system are already included in the *SIMULINK Block Diagram Library*. Some component examples are a transfer function block, general function block, and saturation block and many others. This makes simulation design more attractive and the use of such a package leads an easier understanding of the system than the programming language implementation. Apart from these it is easy to capture, store, process, and display results using the predefined functions. Also the integration algorithm included with the software is accurate and one can also view how signals flow in the Simulation diagram.

The inputs of a squirrel cage induction machine are the three phase voltages, their fundamental frequency, and the load torque. The outputs, on the other hand, are the three phase currents, the electrical torque, and the rotor speed.

FIELD-ORIENTED CONTROL

The control and estimation of ac drives in general are considerably more complex than those of dc drives, and this complexity increases substantially if high performances are demanded. The main reason for this complexity is the need of variable frequency, harmonically optimum converter power supplies, the complex dynamics of ac machines, machine parameter variations, and the difficulties of processing feed back signals in the presence of harmonics.

Even though there are so many strategies have been proposed for induction motor drives; to name a few, slip control, torque control [2, 3, and 4], phase angle control, the very popular methods are scalar control [3] and more over vector control (or) field oriented control[11]. Let us discuss about those two controls.

3.1 SCALAR CONTROL

Scalar control [4], as the name indicates, is due the magnitude variation of the control variables only, and disregards the coupling effect in the machine. For example, the voltage of the machine can be controlled to control the flux, and frequency (or) slip can be controlled to control the torque. However flux and torque are also the functions of frequency and voltage respectively. Scalar controlled drives are some what inferior performance, but they are easy to implement. Scalar controlled drives have been widely used in industry. However, their importance has been diminished recently because of superior performance of vector controlled drives, which is demanded in many applications.

The major disadvantage of the V/Hz method is its sluggish dynamic response since the method disregards the inherent machine coupling. A step change in the speed command produces a slow torque response. During the transient, the magnitude of the stator flux is not maintained (the magnitude decreases) and the machine's torque response is not sufficiently fast. In addition, there is some amount of under damping in the machine's flux and torque responses that increases at lower frequencies. In some operating regions, the system may become unstable.

3.2 FIELD ORIENTED CONTROL

Even though the scalar control is some what easy to implement, but the inherent coupling effect (i.e., both torque and flux are functions voltage or current and frequency) gives sluggish response and the system is easily tends to instability because of higher order (fifth order) system effect [3]. To make it more clear, if, for example, the torque is increased by incrementing slip (i.e., the frequency), the flux tends to decrease. Note that the flux variation is always sluggish. The flux decrease is then compensated by the sluggish flux control loop feeding in additional voltage. This temporary dipping of flux reduces the torque sensitivity with slip and lengthens the response time. This explanation is valid for current fed inverter drives.

The foregoing problems can be solved by vector or field oriented control [1, 5, 6, 8, 10, 12, 13, 14, 24, 26 and 28]. The invention of field oriented control in the beginning of 1970s, and the demonstration that an induction motor can be controlled like a separately excited dc motor, brought a renaissance in the high performance control of ac drives. Because of dc machine like performance [3], vector control is also known as decoupling, orthogonal, or trans vector control. Field oriented control is applicable to both induction motor ad synchronous motor drives. Undoubtedly, vector control and the corresponding feed back signal processing, particularly for modern senseless vector control [1 and 21], are complex and the use of powerful microcomputer or DSP in mandatory. It appears that eventually, field oriented control will oust scalar control, and will be accepted as the industry-standard control ac drives.

3.2.1 Principles of Field-oriented Control

The principle of field oriented control implementation [3,11] can be explained with the help of Fig.3.1 [3] where the machine model is represented in synchronously reference frame. The inverter is omitted from the figure, assuming that it has unity current gain, that is, it generates currents i_a, i_b, i_c as dictated by the corresponding command currents i_a^*, i_b^* and i_c^* from the controller.

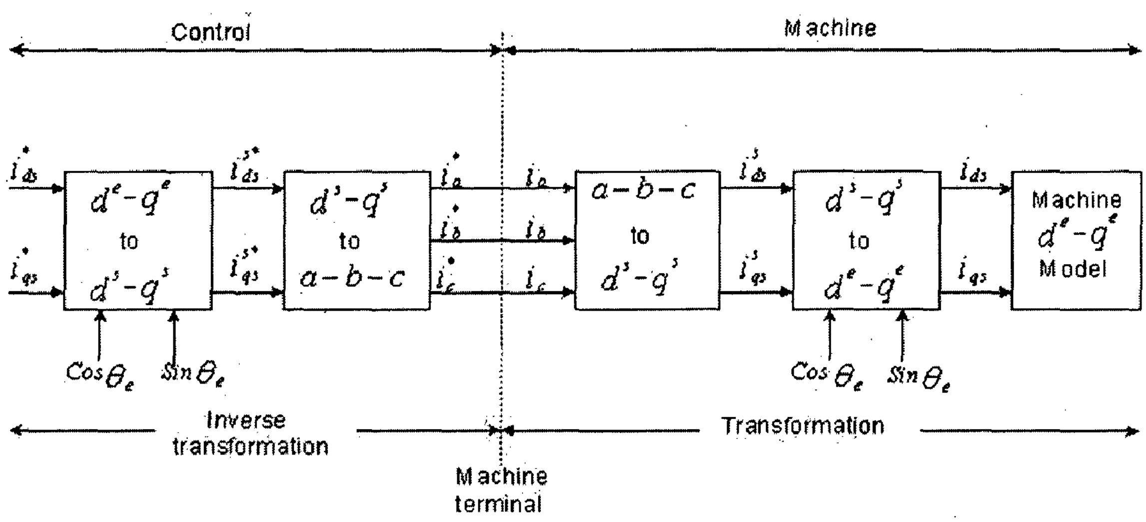

Fig. 3.1 Field orientation principle with machine in synchronously rotating frame model

The machine terminal phase currents i_a, i_b, i_c are converted to i_{ds}^s and i_{qs}^s components by the $3\phi/2\phi$ transformation. These are then converted to synchronously rotating reference frame by the unit vector components cos (θ) and sin (θ) before applying them to the d-q model of induction motor. Then the controller makes two stages of inverse transformation, so that the control currents i_{ds}^{s*} and i_{qs}^{s*} correspond to the machine currents i_{ds}^s and i_{qs}^s, respectively. In addition unit vector assures correct alignment of i_{ds}^s current with the flux vector and i_{qs}^s perpendicular to it.

3.3 CLASSIFICATION OF FIELD-ORIENTED CONTROL METHOD

Field-oriented control methods are classified by how the unit vector (cos (θ) and sin (θ)) is generated for the control. There are essentially two general methods of field oriented control [3, 11 and 39].

1. Direct field-oriented control
2. Indirect field-oriented control

3.3.1 Indirect Field-Oriented Control

Hasse [28] invented the indirect filed-oriented or feed forward control method. The indirect field-oriented control method is essentially same as the direct field-oriented control method, except the unit vector signals are generated in feed forward manner. Fig.3.2 explains the fundamental principle of indirect field-oriented control with the help of phasor diagram [1, 11]. The d^s-q^s axes are fixed on the stator, but the d^r-q^r axes, which are fixed on the rotor, are moving at speed ω_r as shown. Synchronously rotating axes d^e-q^e are rotating ahead of the d^r-q^r axes by the positive slip angle θ_{sl} corresponding to slip frequency ω_{sl}. Since the rotor pole is directed on the d^e axis and $\omega_e = \omega_{sl} + \omega_r$, we can write

$$\theta_e = \int \omega_e dt = \int (\omega_r + \omega_{sl}) dt = \theta_r + \theta_{sl} \tag{3.1}$$

Note that the rotor position is not absolute, but is slipping with respect to the rotor at frequency ω_{sl}. The phasor diagram suggests that for decoupling control, the stator flux component of current i_{ds} should be aligned on the d^e-axis, and the torque component of current i_{qs} should be on the q^e-axis, as shown.

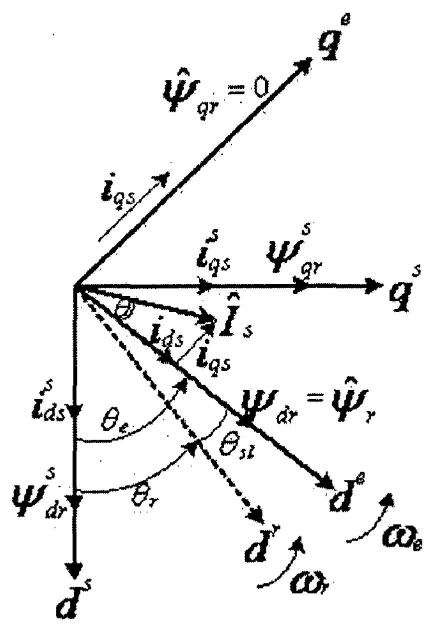

Fig.3.2 Phasor diagram for Indirect vector control

The overall block diagram of the indirect vector control [3, 11] is shown in Fig.4.2. In the indirect vector control torque can be controlled by regulating i_{qs}^{e} and slip speed, rotor flux can be controlled by regulating i_{ds}^{e}.

For the desired value of T_{em} at the given level of rotor flux, the desired value of i_{qs}^{e} may be obtained from the torque equation in terms of rotor flux and stator current. A rotor slip calculation is used to find the slip speed that is integrated to give the slip position. Adding this to the rotor position measurement gives the rotor flux position and hence we can obtain the unit vectors required to transform between the stationary frame and rotating frame quantities.

This slip calculator requires the correct T_r, which is used to calculate the slip command to establish the rotor flux angle with respect to the rotor axis. Variations of the T_r are mainly caused by the thermal drift of the rotor resistance and by the change of the rotor inductance due to saturation. If the parameters of the machine and controller are not identical, it could lead to the saturation of the machine by the wrong slip command i.e., the position of the flux in the in direct vector control is not aligned with the real flux axis.

3.3.2 Limitations of in-direct field-orientation control

Vector control transforms the control of an induction motor to that of a separately excited dc motor by creating independent channels for flux and torque control. Crucial to the success of the vector control scheme is the knowledge of the instantaneous position of the rotor flux. The position of the rotor flux is measured in the direct vector control scheme and estimated in the indirect vector control scheme. This requires a priori knowledge of the machine parameters which makes the indirect vector control scheme machine parameter dependent. Changes in temperature and saturation levels of the machine vary the steady state and dynamic operation of the drive system.

The rotor resistance variation with temperature and also the saturation highly affects all calculations. Hence in high performance drives using indirect field orientation control must be faced at least two challenges [3, 4 and 11]:

- Elimination of any mechanical equipment for speed or angle detection leading to sensor less vector control.
- Identification of electrical and mechanical parameters of the drive leading to adaptive control.

3.4 FUZZY LOGIC CONTROLLER

The main feature of fuzzy logic controllers (FLCs) is that linguistic, imprecise knowledge of human experts is used. However, the implementation of conventional fuzzy logic controllers suffers from the disadvantage that no formal procedures exist for the direct incorporation of the expert knowledge during the development of the controller. The structure of the fuzzy controller (number of rules, the rules themselves, number and shape of membership functions, etc) is achieved through a time consuming tuning process which is essentially manual in nature. The ability to automatically 'learn' characteristics and structure which may be obscure to the human observer is, however, inherent in neural networks. A fuzzy logic-type controller having the structure of a neural network offers the advantages of both - the ability of fuzzy logic to use expert human knowledge and the learning ability of the neural network - and overcomes their disadvantages - the lack of a formal learning procedure for the fuzzy controller fuzzy controller and the lack of a clear correlation with the physical problem when using neural networks.

The conventional controllers for vector controlled induction motor drive (VCIMD) suffer from the problem of stability, besides these controllers show either steady state error or sluggish response to the perturbation in reference setting or during load perturbation. The motor control issues are traditionally handled by fixed gain PI and Proportional-integral-derivative (PID) controllers. However, the fixed-gain controllers are very sensitive to parameter variations, load disturbances, etc. Thus, the controller parameters have to be continually adapted. However, it is often difficult to develop an accurate system mathematical model due to unknown load variation, temperature variations, unknown and unavoidable parameter variations due to saturation and system disturbances. In order to overcome the above problems, recently, the fuzzy-logic controller (FLC) is being used for motor control purpose. The mathematical tool for the

FLC is the fuzzy set theory introduced by Zadeh [28]. As compared to the conventional PI, PID, and their adaptive versions, the FLC has some advantages such as:

1. It does not need any exact system mathematical model.
2. It can handle nonlinearity of arbitrary complexity.
3. It is based on the linguistic rules with an IF-THEN general structure, which is the basis of human logic.

However, the application of FLC has faced some disadvantages during hardware and software implementation due to its high computational burden.

3.4.1 Modeling of Fuzzy Logic Based Controller

Fuzzy variables and control rules:

In order to obtain better control results, it is necessary to use appropriate number of fuzzy variables and to formulate appropriate control rules. In this study, we use the fundamental seven kinds of fuzzy variables as follows:

NL : Negative Large — PL : Positive Large

NM : Negative Medium — PM : Positive Medium

NS : Negative Small — PS : Positive Small

ZE : Approximately Zero

The next step is to decide the appropriate shape of the membership functions for w_e and w_{ee}. More fuzzy sets in w_e and w_{ee} will lead to higher precision in this input space. Hence seven fuzzy sets are assigned to each of the inputs in their respective universe of discourse. Consequently, if the number of fuzzy sets in a particular universe of discourse is increased to infinity, then all fuzziness will be lost and it will be equivalent to a conventional input domain. To simplify mathematical computations, the shape of the fuzzy sets on the two extreme ends of the respective universe of discourse is taken as trapezoidal whereas all other intermediate fuzzy sets are triangular. This is illustrated in fig.3.4.

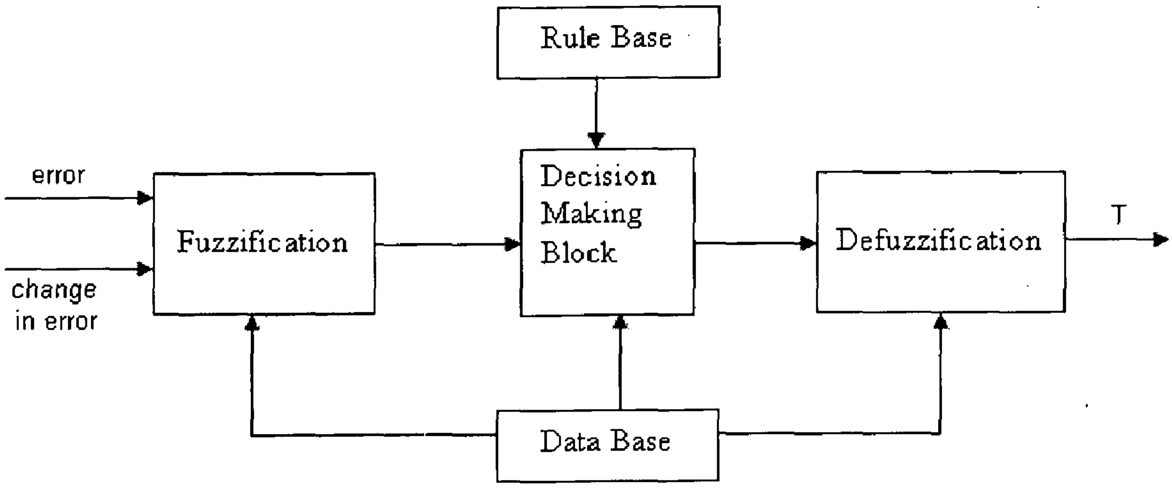

Figure 3.3 Block diagram of Fuzzy Logic Controller (FLC)

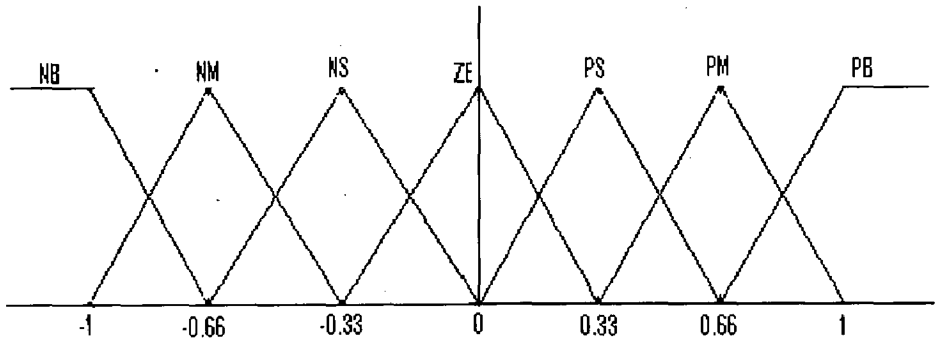

Figure 3.4 Fuzzy sets considered for control

The control rules for the FLC can be described by language using the input variables w_e and w_{ee}, and the output variable, T. For example the i-th control rule can be usually written as:

Rule i : if w_e is F_i and w_{ee} is G_i then T is H_i.

Where F_i, G_i and H_i are fuzzy variables.

In general, it is difficult to formulate control rules for an unknown system. However, we already know the system and can predict a step response of the motor speed. Therefore it is comparatively easy to formulate control rules.

The typical step response of the speed from 0 rpm to a set value as shown in fig.3.5

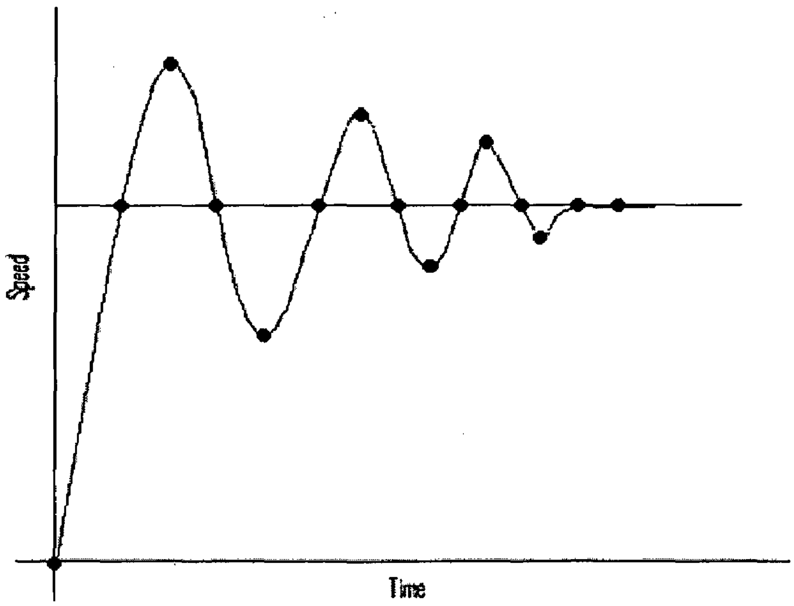

Figure 3.5. General Step Response of the speed

The characteristic points are shown with dots in this figure. To formulate control rules, it is necessary to examine the condition at each characteristic point and to consider the relation among E_r, E_r (dot) and ΔT_c, so as to bring the step response close to the set speed value. The fuzzy rule table used in this work is given in table 3.1 [27].

e	**NB**	**NM**	**NS**	**ZE**	**PS**	**PM**	**PL**
NB	NVB	NVB	NB	NM	NS	NVS	ZE
NM	NVB	NB	NM	NS	NVS	ZE	PVS
NS	NB	NM	NS	NVS	ZE	PVS	PS
ZE	NM	NS	NVS	ZE	PVS	PS	PM
PS	NS	NVS	ZE	PVS	PS	PM	PL
PM	NVS	ZE	PVS	PS	PM	PL	PVB
PL	ZE	PVS	PS	PM	PL	PVB	PVB

Table 3.1 Logic rules for Fuzzy Logic (FL) controller

Fuzzy controllers are very simple conceptually. They consist of an input stage, a processing stage, and an output stage. The input stage maps sensor or other inputs, such as switches, thumbwheels, and so on, to the appropriate membership functions and truth values. The processing stage invokes each appropriate rule and generates a result for each, then combines the results of the rules. Finally, the output stage converts the combined result back into a specific control output value.

The most common shape of membership functions is triangular, although trapezoids and bell curves are also used, but the shape is generally less important than the number of curves and their placement. From three to seven curves are generally appropriate to cover the required range of an input value, or the "universe of discourse" in fuzzy jargon.

In practice, the fuzzy rule sets usually have several antecedents that are combined using fuzzy operators, such as AND, OR, and NOT, though again the definitions tend to vary: AND, in one popular definition, simply uses the minimum weight of all the antecedents, while OR uses the maximum value. There is also a NOT operator that subtracts a membership function from 1 to give the "complementary" function.

There are several different ways to define the result of a rule, but one of the most common and simplest is the "max-min" inference method, in which the output membership function is given the truth value generated by the premise.

Rules can be solved in parallel in hardware, or sequentially in software. The results of all the rules that have fired are "defuzzified" to a crisp value by one of several methods. There are dozens in theory, each with various advantages and drawbacks.

The "centroid" method is very popular, in which the "center of mass" of the result provides the crisp value. Another approach is the "height" method, which takes the value of the biggest contributor. The centroid method favors the rule with the output of greatest area, while the height method obviously favors the rule with the greatest output value.

Fuzzy control system design is based on empirical methods, basically a methodical approach to trial-and-error. The general process is as follows:

- Document the system's operational specifications and inputs and outputs.
- Document the fuzzy sets for the inputs.
- Document the rule set.
- Determine the defuzzification method.
- Run through test suite to validate system, adjust details as required.
- Complete document and release to production.

CHAPTER 4

ADAPTIVE OBSERVER BASED CONTROL OF INDUCTION MOTOR

4.1 INTRODUCTION

To reduce total hardware complexity and costs and to increase mechanical robustness, it is desirable to eliminate speed and position sensors in vector- controlled drives [2, 3]. In drives operating in hostile environments or in high-speed drives, speed sensors cannot be mounted. As real-time computation costs are continuously decreasing, speed estimation can be performed by using various software-based estimation techniques where stator voltages and/or currents are monitored on-line. Sometimes it is possible to use a scheme where the dc link voltage is monitored and this information is combined with information on the switching states of the inverter. It is also possible to use other types of solutions, e.g. the stator phase third harmonic methods which are based on direct stator flux estimation utilizing monitored stator voltages and currents yield unsatisfactory performance at the low speed region, since they use pure integration for the stator flux and they are sensitive to the offset voltage of the voltage sensors and the variation of the stator resistance.

The most significant limitation of speed and flux estimation methods that rely on back emf is lack of robustness at low to zero speed. Techniques which include the use of MRAC systems are all limited by the fact that the back emf becomes nearly zero at very low speeds and at low frequency the integration of the voltages is problematic. Incorporation of the mechanical system model including torque estimate feed forward into the MRAC significantly improves the speed estimation dynamics, including low speed transients. Incorporation of a closed-loop observer into the MRAC improves zero speed operation.

There is a strong industrial need for the development and exploitation of systems incorporating artificial- intelligence-based soft computing controllers because of the numerous advantages offered,

4.2 SPEED ESTIMATION

There are various possibilities to obtain open-loop flux and speed estimators. For this purpose it is possible to use monitored terminal voltages and currents, or monitored currents together with the monitored dc link voltage. If the latter strategy is used then knowledge on the switching states of the inverter can be employed to reconstruct the voltages. The most direct and simple way to determine the stator and rotor fluxes is to utilize monitored stator currents and monitored or reconstructed stator voltages. The stator flux linkages can then be obtained by integration.

The stator resistance is also required. The rotor fluxes can then be obtained using additional machine parameters, e.g. the stator transient inductance and the ratio of the magnetizing inductance to the inductance of the rotor. However, at low speed there are well-known problems, since the ohmic drop is large and even small measurement errors and an imprecise value of the stator resistance can lead to significant errors. Even at high speed it is important to have an accurate representation of the stator resistance. In a drive containing a speed sensor, the low speed problems are usually avoided by using a flux mode which relies on the rotor voltage equations, and utilizes the well-known slip relation.

This technique is sensitive to the rotor time constant and, at high speed to speed measurement errors. To avoid some of these problems, in a drive with a speed-sensor, it is possible to use a hybrid flux estimator. In this case at low speed the rotor-voltage equation-based flux model is used and for high-speed the stator voltage equation is utilized.

It is also possible to obtain real-time estimates of the speed by utilizing various flux linkages in the machine. This is based on the physical fact that the rotor speed is the difference between the speed of the flux vector considered and the slip speed. The flux vector speed can be obtained form the terminal voltages and currents, by using the components of the flux vector (which can be obtained by the techniques described above). The slip speed can be obtained by also using the torque producing current component. In a scheme where the slip frequency is also a function of the derivative of the torque producing current, the sensitivity to noise is an important factor. The techniques utilizing the flux linkages for speed determination have similar problems to those associated with the flux estimators described above. One important problem is parameter sensitivity, but this depends on the choice of the estimated flux.

In a laboratory implementation, low-pass filters are used to remove high frequency voltage components. Signal offset at integrator inputs must also be removed to prevent saturation. For this purpose, during calibration of the voltage and current sensors an average offset is obtained and this is subtracted from the measured value during operation. The integration step length has also an important effect on the performance of the estimator.

It should be noted that there are inherent errors in any implementation of the flux estimator and these errors will have a detrimental effect on the drive performance. However, if fuzzy controllers are used, an improvement in the overall drive performance can be expected due to the tolerance of these controller types to imprecision. An alternative approach is to reduce the Error in the flux estimator employing by extended observers which are inherently closed loop in nature.

To obtain the full-order state adaptive observer, first the model of the induction machine is considered in the stationary reference frame and then an error compensator term is added to this.

4.3 PROBLEMS WITH ESTIMATION [36]

Before looking into individual approaches, the common problems of the speed and flux estimation are discussed briefly for general field-orientation [36].

4.3.1 Parameter sensitivity

One of the important problems of the sensorless control algorithms for the sensorless IM drives is the insufficient information about the machine parameters which yield the estimation of some machine parameters along with the sensorless structure. Among these parameters, stator resistance, rotor resistance and rotor time-constant play more important role than the other parameters since these values are more sensitive to temperature changes.

The knowledge of the correct stator resistance R_s is important to widen the operation region toward the lower speed range. Since at low speeds the induced voltage is low and stator resistance voltage drop becomes dominant, a mismatching stator resistance induces instability in the system. On the other hand, errors made in determining the actual value of the rotor resistance R_r may cause both instability of the system and speed estimation error proportional to R_r. Also, correct T_r value is vital decoupling factor in the sensor less control scheme.

4.3.2 Pure integration

The other important issue regarding many of the topologies is the integration process inherited from the IM dynamics where an integration process is needed to calculate the state variables of the system. However, it is difficult both to decide on the initial value, and prevent the drift of the output of a pure integrator. Usually, to overcome this problem a low-pass filter replaces the integrator.

4.3.3 Overlapping-loop problem

In a sensor less control system, the control loop and the speed estimation loop may overlap and these loops influence each other. As a result, outputs of both of these loops may not be designed independently and in some bad cases this dependency may influence the stability or performance of the overall system. The algorithms, where terminal quantities of the machine are used to estimate the fluxes and speed of the machine, are categorized in two basic groups. First one is "the open-loop observers" in a sense that the on-line model of the machine does not use the feedback correction. Second one is "the closed-loop observers" where the feedback correction is used along with the machine model itself to improve the estimation accuracy.

4.4 FULL ORDER ADAPTIVE STATE OBSERVER

An Estimator is a dynamic system whose state variables are estimates of some other system (e.g. electrical machine).There are basically two forms of the implementation of an estimator: open-loop and closed-loop, the distinction between two being whether or not a correction term, involving the estimation error, is used to adjust the response of the estimator. A closed loop estimator is referred to as an observer[2].

In this dissertation a full order adaptive state observer is implemented. It's a modified estimator which can be used to estimate the rotor flux linkages of an induction machine. And is then modified so it can also yield the speed estimate, and thus an adaptive speed estimator is derived from that estimator. To obtain a stable system, the adaptation mechanism is derived by using the state-error dynamic equations together with lyaponav's stability theorem. In an inverter-fed drive system, the observer uses the monitored stator currents together with the monitored stator voltages or reference stator voltages.

4.5 STATE OBSERVER

A state observer estimates the state variables based on the measurements of the output and control variables. Consider the system defined by

$$\dot{x} = Ax + Bu \tag{4.1}$$

$$y = Cx \tag{4.2}$$

Assume that the state x is to be approximated by the state $\tilde{x}$ of the dynamic model

$$\dot{\tilde{x}} = A\tilde{x} + Bu + G(y - C\tilde{x}) \tag{4.3}$$

which represents the state observer. It can be seen that the observer has y and u as inputs and $\dot{\tilde{x}}$ as output. The last term on the right-hand side of this model equation, (4.3), is a correction term that involves the difference between the measured output y and the estimated output $C\tilde{x}$. Matrix G serves as a weighting matrix. The correction term monitors the state $\tilde{x}$. In the presence of discrepancies between the A and B matrices used in the model and those of the actual system, the addition of the correction term, will help to reduce the effects due to difference between the dynamic model and the actual system. Fig 4.1 shows the block diagram of the system and the full-order state observer.

To obtain the observer error equation, subtract equation (4.3) from equation (4.1)

$$\dot{x} - \dot{\tilde{x}} = Ax - A\tilde{x} - G(Cx - C\tilde{x})$$

$$= (A - GC)\,(x - \tilde{x}) \tag{4.4}$$

Define the difference between x and $\tilde{x}$ as the error vector e, or

$$e = x - \tilde{x} \tag{4.5}$$

then (4.4) becomes

$$\dot{e} = (A - GC)\,e \tag{4.6}$$

We see that the dynamic behavior of the error vector is determined by the eigenvalues of matrix $(A - GC)$. G is chosen in such a way that the dynamic behavior of the error vector is asymptotically stable and is adequately fast, and then the error vector will tend to zero (origin) with an adequate speed.

If the system is completely observable, then it can be proved that it is possible to choose matrix G such that $(A - GC)'$ has arbitrarily desired eigenvalues. That is, the observer gain matrix G can be determined to yield the desired matrix $(A - GC)$.

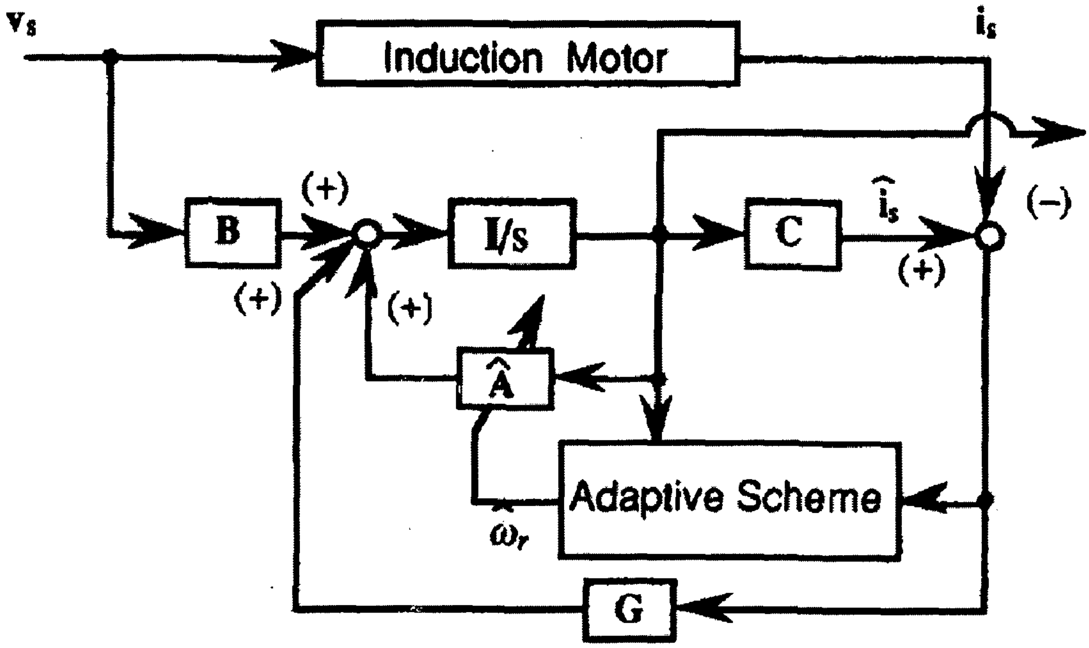

Fig.4.1.Full-order state Adaptive speed observer

4.6 OBSERVER MODEL [2]

The dynamic model of induction motor can be written choosing as state variables currents and/or fluxes for the electrical equations, and the rotor speed for the mechanical equation.

Assuming the rotor speed to be measurable, ω_r can be considered as a time varying parameter so that the dynamic model can be rewritten as the following fourth order linear time varying system:

$$\dot{x} = A(w_r)x + Bu \tag{4.7}$$

where

$$x = \left(i_D, i_Q, \Phi_d, \Phi_q\right)^T \tag{4.8}$$

$$u = \left(v_D, v_Q\right)^T \qquad (4.9)$$

Where u is the space vector of stator voltages.

and the system matrices are

$$A(w_r) = \begin{pmatrix} a11 & a12 \\ a21 & a22 \end{pmatrix} \qquad (4.10)$$

$$B = \begin{pmatrix} \frac{1}{\sigma Ls} & 0 & 0 & 0 \\ 0 & \frac{1}{\sigma Ls} & 0 & 0 \end{pmatrix}^T . \qquad (4.11)$$

with

A11 = -(Rs + Lm^2*Rr/Lr^2) / (σ *Ls) * I (4.12)

A12 = Lm / (σ *tr*Ls*Lr) * I; (4.13)

A21 = Lm/tr * I; (4.14)

A22= -1/tr * I; (4.15)

Aw = [Lm/(σ *Ls*Lr) * I; -I]; (4.16)

Where Lm and Lr are the magnetizing inductance and rotor self-Inductance respectively.

Assuming that, among electrical state variables, only stator currents are measurable. Then coherent choice for the output variables is $y = (i_D, i_Q)^T$. The dynamic model can be so completed with the output equation

$$y = Cx \qquad (4.17)$$

Where the matrix C is given by

$$C = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{pmatrix}$$

This dynamic model is used for the estimation of rotor speed of the induction motor. The observer can be written as

$$\frac{d\tilde{i}s}{dt} = A_{11}\tilde{i}s + A_{12}\tilde{\phi}_r + \frac{1}{\sigma L_s}v_s + G(\tilde{i}s - i_s) \qquad (4.18)$$

Where the observer gain matrix G is calculated based on the pole placement technique. The selection of the observer pole is based on the compromise between the rapidity of error and the sensitivity to the disturbances and measurement noises.

4.7 MODELLING OF THE DRIVE SYSTEM

The system block diagram consists of the observer, the control structure of the drive, the inverter and an induction motor. All these are modeled separately and are interconnected through various control signals.

4.7.1 Modeling of the vector control of Drive

The vector controller subsystem, computes the reference value of the torque and flux producing components of the stator current. The actual motor speed ω_r and the reference speed command ω^*_r are the inputs to the subsystem. This subsystem compares the reference speed command with the actual speed output, which results in a speed error $\omega_e(t)$.

$$\omega_e(t) = \omega^*_r(t) - \omega_r(t) \qquad (4.19)$$

The resultant speed error is processed in a PI controller, the outputs of the speed controller sets the reference value of torque producing components of the motor stator currents given by

$$i_{qs}^{*} = K_p \cdot \omega_e(t) + K_i \cdot \int \omega_e(t) \quad (4.20)$$

Where, K_p, K_i are proportional and integral gain constants respectively of the speed controller. The value of K_p and K_i depends on the drive system parameters.

The reference flux current component is computes by eq. (4.21),

$$i_{ds}^{*} = \psi_r / L_m \quad (4.21)$$

The subsystem further processed the reference flux current component i_{ds}^{*} and the reference torque current component i_{qs}^{*} through separate PI controller to produce reference d-axis voltage v_{ds}^{*} and reference q-axis voltage v_{qs}^{*} respectively. Here the quantities v_{ds}^{*} and v_{qs}^{*} are reference quantities corresponding to the direct and quadrature component of the stator voltage which should be applied to stator to produce reference current vectors expressed in the synchronously rotating reference frame of d-q variables as in eqs.(4.20) and (4.21). The reference dq-axis component of stator voltage is converted to 3-phase voltage by syn to abc subsystem, for the PWM generator where this reference 3-phase voltage is compared with the carrier triangular wave to generate the switching pulses for the inverter.

The reference dq-axis voltage is converted to 3-phase voltage by syn to abc subsystem. In order to accomplice the task of vector rotation, the system requires a further processing of signals in order to obtain the instantaneous value of the vector rotator $e^{j\theta_e}$. For this purpose, the reference slip speed ω_{sl}^{*}, is added to the sensed rotor speed ω_r and the sum is integrated to obtain the flux angle θ_e. Thus, the vector rotator $e^{j\theta_e}$ is obtained. These voltages v_{ds}^{*} and v_{qs}^{*} are expressed in the synchronously rotating reference frame, therefore are dc in nature. In order to transform them in ac quantities, these voltages are converted in stationary reference frame. For this purpose reverse park transform is used.

$$\begin{bmatrix} v_{qs}^{s*} \\ v_{ds}^{s*} \end{bmatrix} = \begin{bmatrix} \cos\theta_e & \sin\theta_e \\ -\sin\theta_e & \cos\theta_e \end{bmatrix} \begin{bmatrix} v_{qs}^{*} \\ v_{ds}^{*} \end{bmatrix} \quad (4.22)$$

Therefore magnitude and phase angle of the signals v_{ds}^{*} and v_{qs}^{*} are controlled instantaneously thereby fulfilling basic requirement of vector control.

The reference voltages v_a^*, v_b^* and v_c^* are obtained with the help of following mathematical equations.

$$\begin{bmatrix} v_a^* \\ v_b^* \\ v_c^* \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ \sqrt{\frac{3}{2}} & -\frac{1}{2} \\ \sqrt{\frac{3}{2}} & -\frac{1}{2} \end{bmatrix} \begin{bmatrix} v_{ds}^{s*} \\ v_{qs}^{s*} \end{bmatrix} \tag{4.23}$$

These three voltages v_a^*, v_b^* and v_c^* which are generated from vector controlled block are voltages should be maintained across the winding of the motor for the purpose of its vector control. On maintaining these voltages, the instantaneous control of magnitude and phase of the two components of primary current vector is affected and thereby, the requirement of vector control is fulfilled.

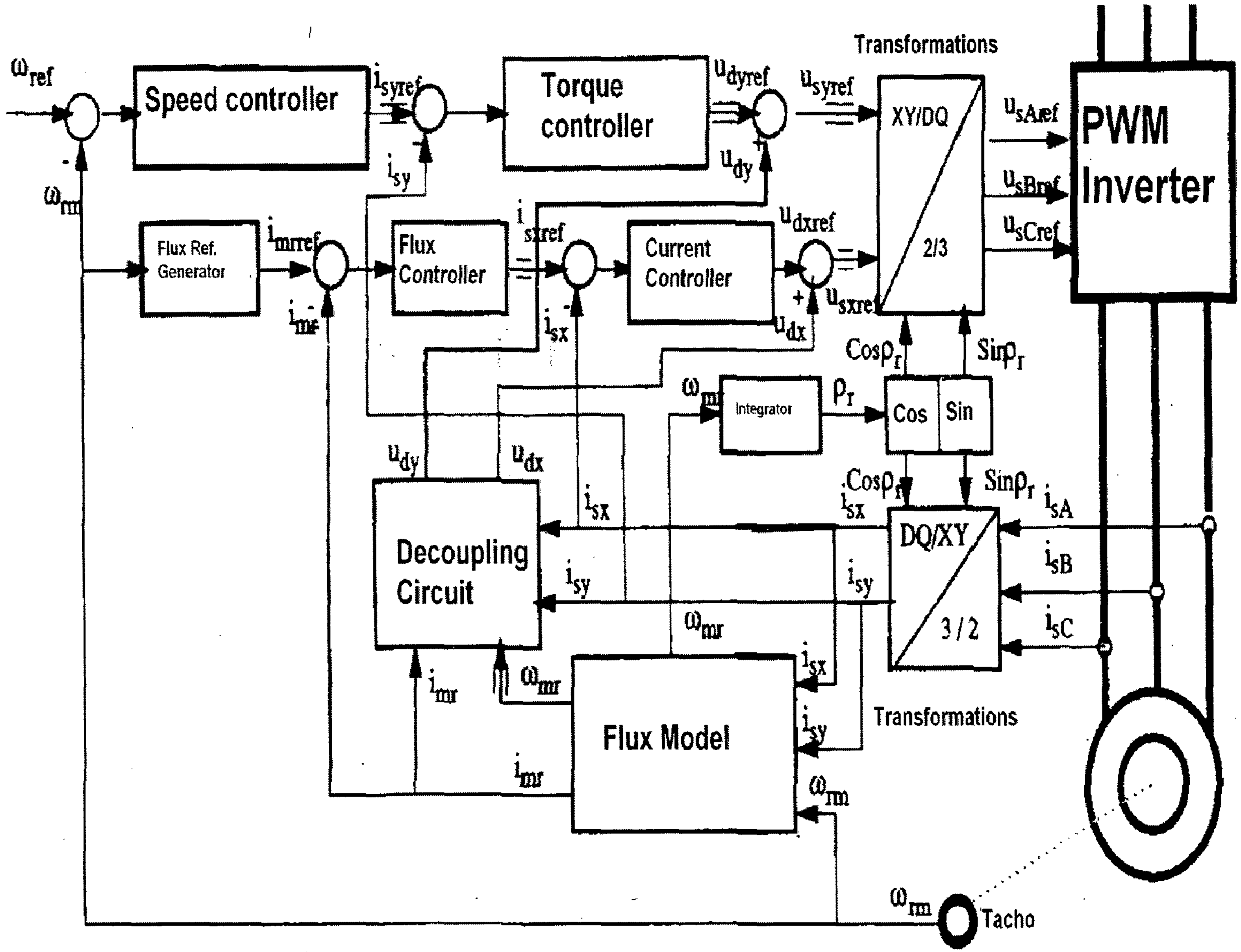

Fig. 4.2 Schematic of the Indirect rotor flux orientedcontrol of induction machine

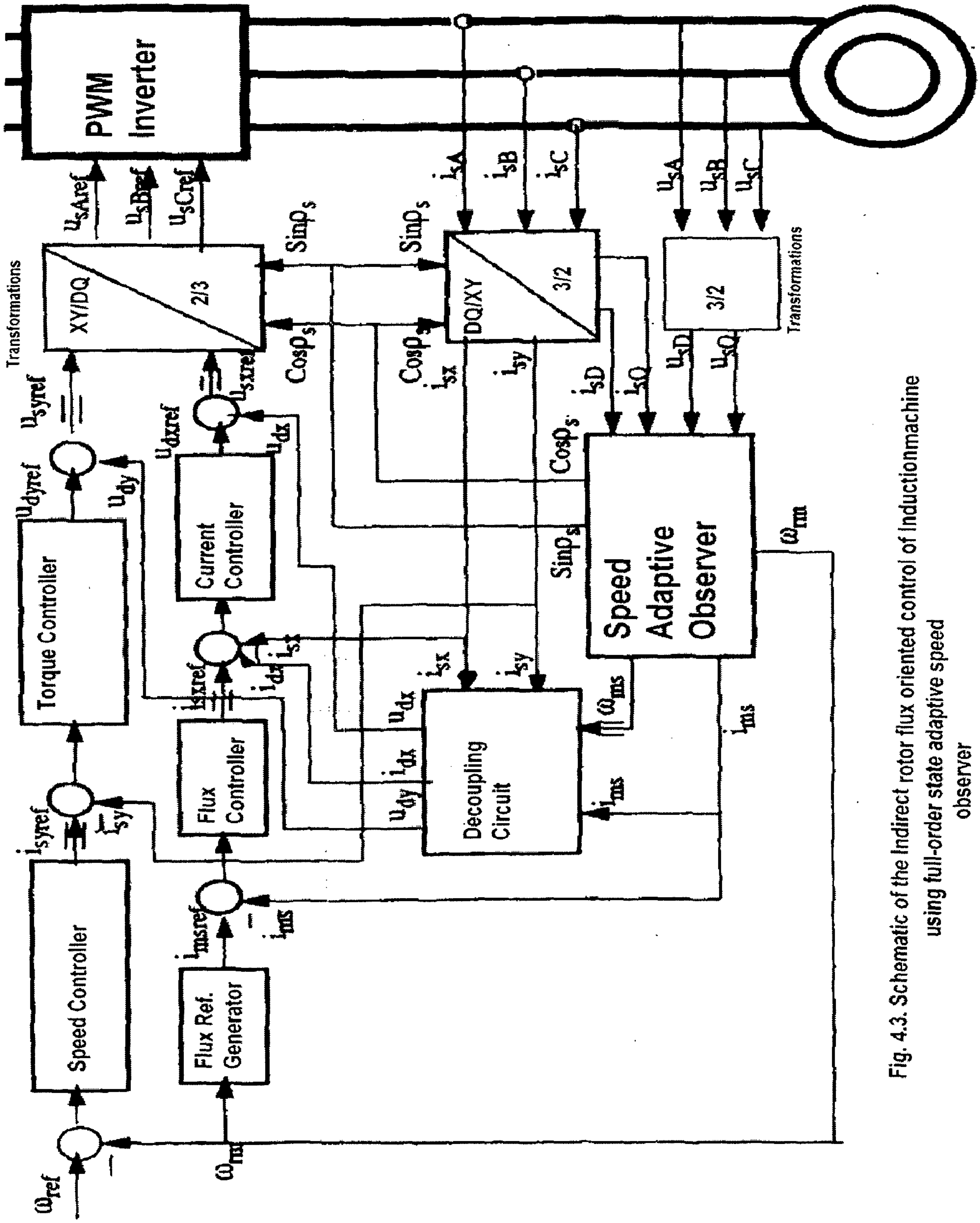

Fig. 4.3. Schematic of the Indirect rotor flux oriented control of Inductionmachine using full-order state adaptive speed observer

CHAPTER 5

RESULTS AND DISCUSSIONS

Simulation of the proposed MATLAB design is shown in the fig (4.2). The induction motor used in drive system is simulated in the MATLAB/SIMULINK with the dynamic space phasor model using nominal parameters given in appendixA. The behavior of machine has been observed under free acceleration and rated load condition with voltage is rated rms line-to-line voltage. At the stall the input impedance of the motor essentially the stator resistance and leakage reactance. Consequently with the rated voltage applied, the starting current is large.

5.1 VECTOR CONTROL WITH SENSOR

5.1.1. Performance of the 1HP Vector controlled Induction motor drive

The simulation results of the 1 HP Vector controlled Induction motor drive are presented in the figures from 5.1 to 5.6 for the operation of the drive for different speeds and loads by using Proportional and Integral (PI)controller and Fuzzy logic controller.

Fig.5.1 gives the speed response of induction motor drive ffor different loads set by the step source externally. The starting speed reference is 2000rpm.The speed of the drive reachesthe reference speed with in 1secand after that speed reference is changed from 2000 rpm to 1000 rpm. From the Fig.it is observed that speed reaches its reference speed within 0.1 sec.Again the speed reference changed to 1000 rpm and 500 rpm.

The electromagnetic torque produced is shown in Fig 5.2. It settles the load torque which changing from noload.to rated load . The response of the drive is fast. One can observe the sudden variations in the torque while changing the speed from one value to the other.Fig.5.3 shows the three phase winding currents and they are approximately sinusoidal. Fig 5.5 and 5.6 shows the speed and torque response of the drive when Fuzzy controller is used instead of PI controller

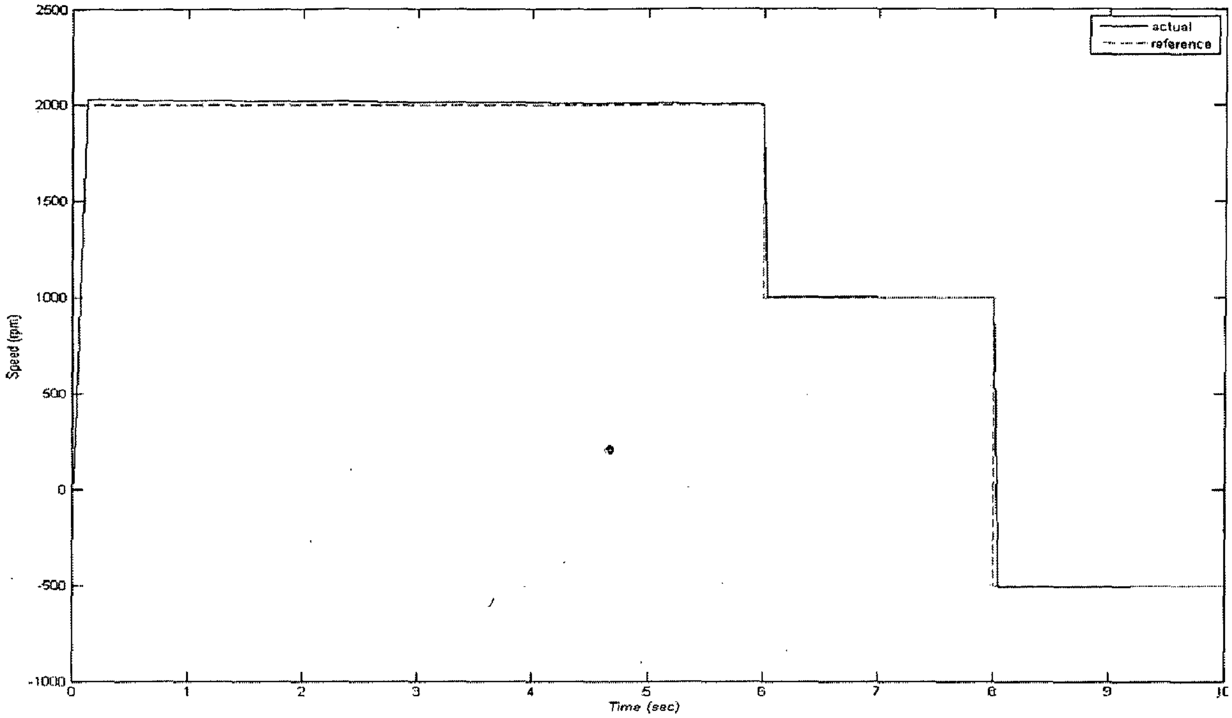

Fig.5.1. Speed Response with PI controller

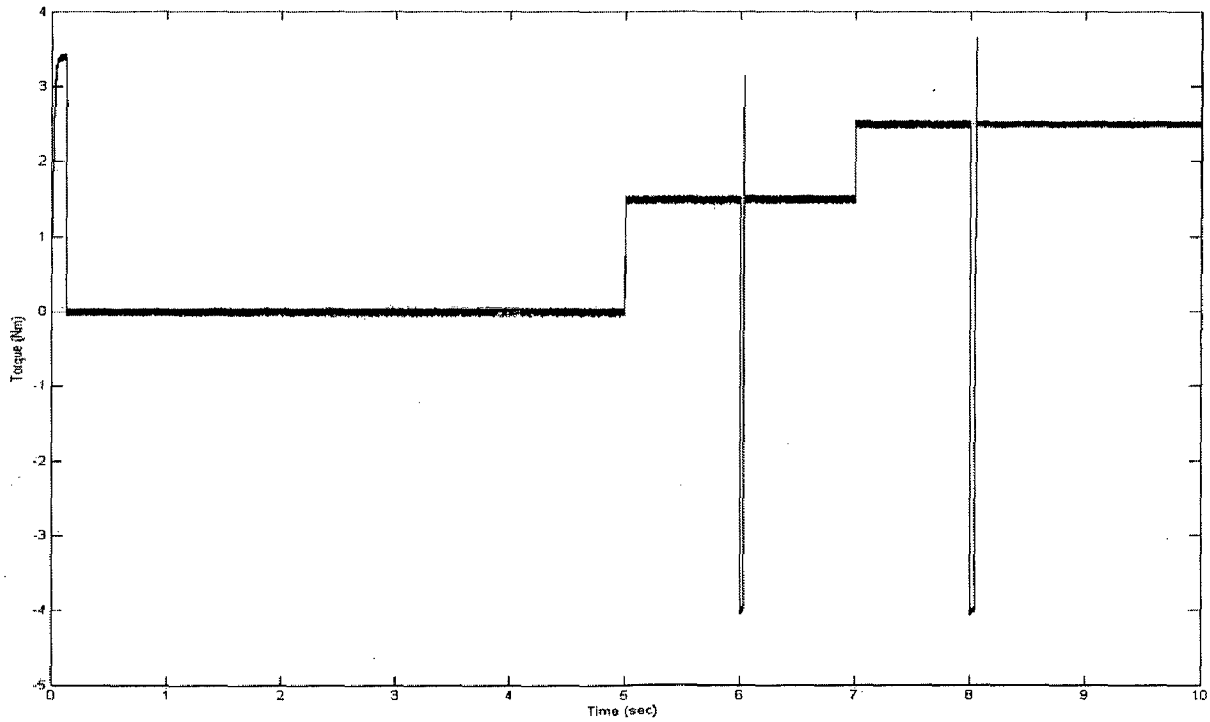

Fig.5.2.Torque response with PI controller

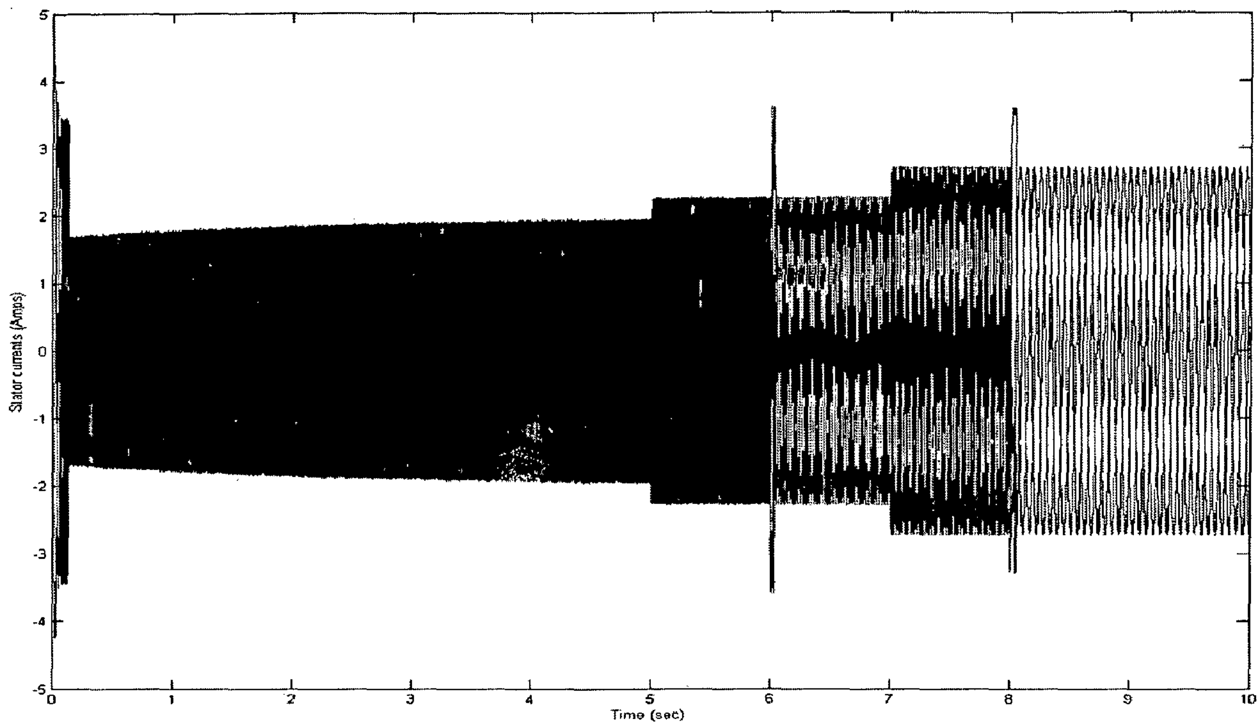

Fig.5.3. Stator currents response with PI Controller

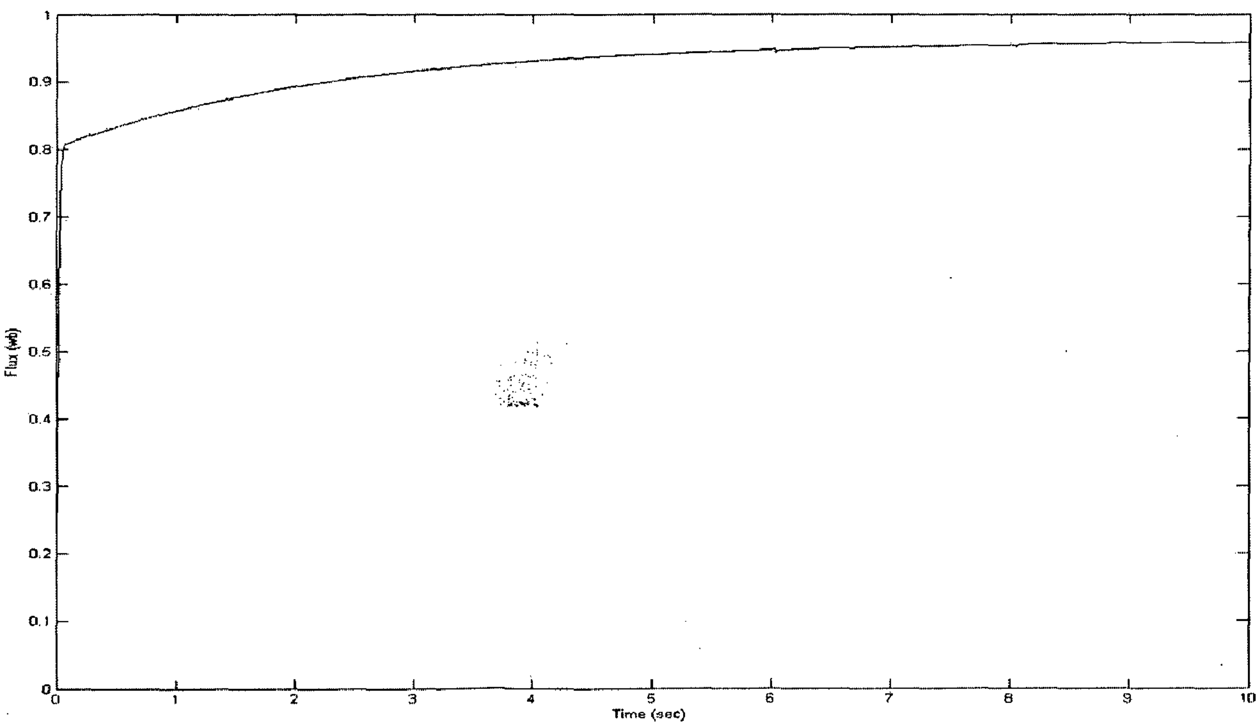

Fig.5.4. Flux response

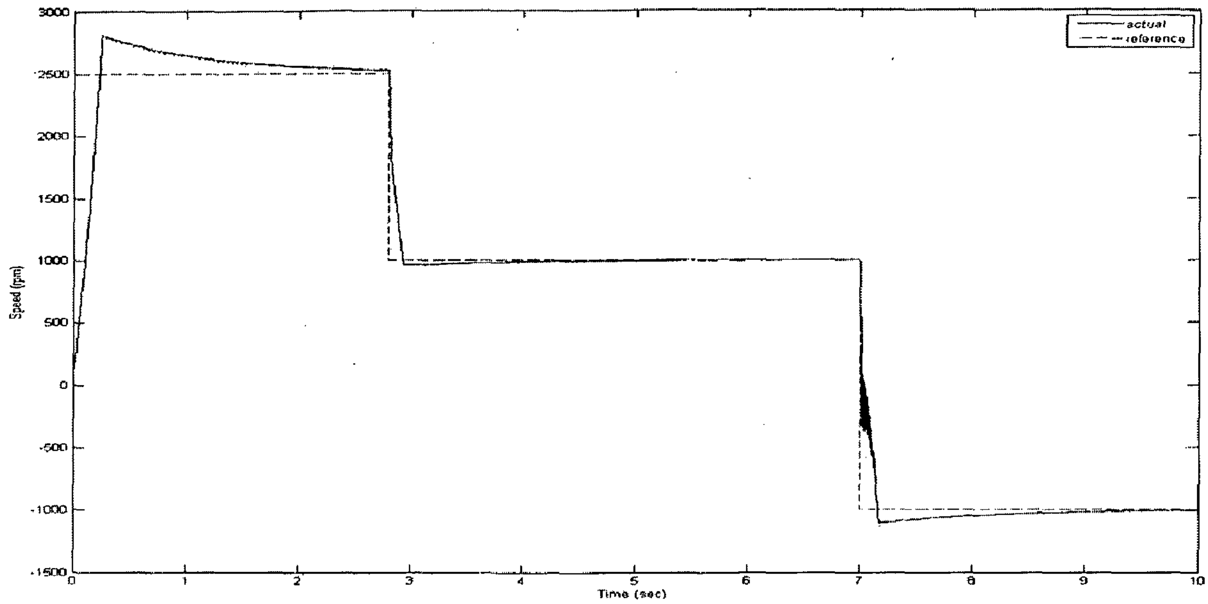

Fig.5.5 Speed response with fuzzy controller

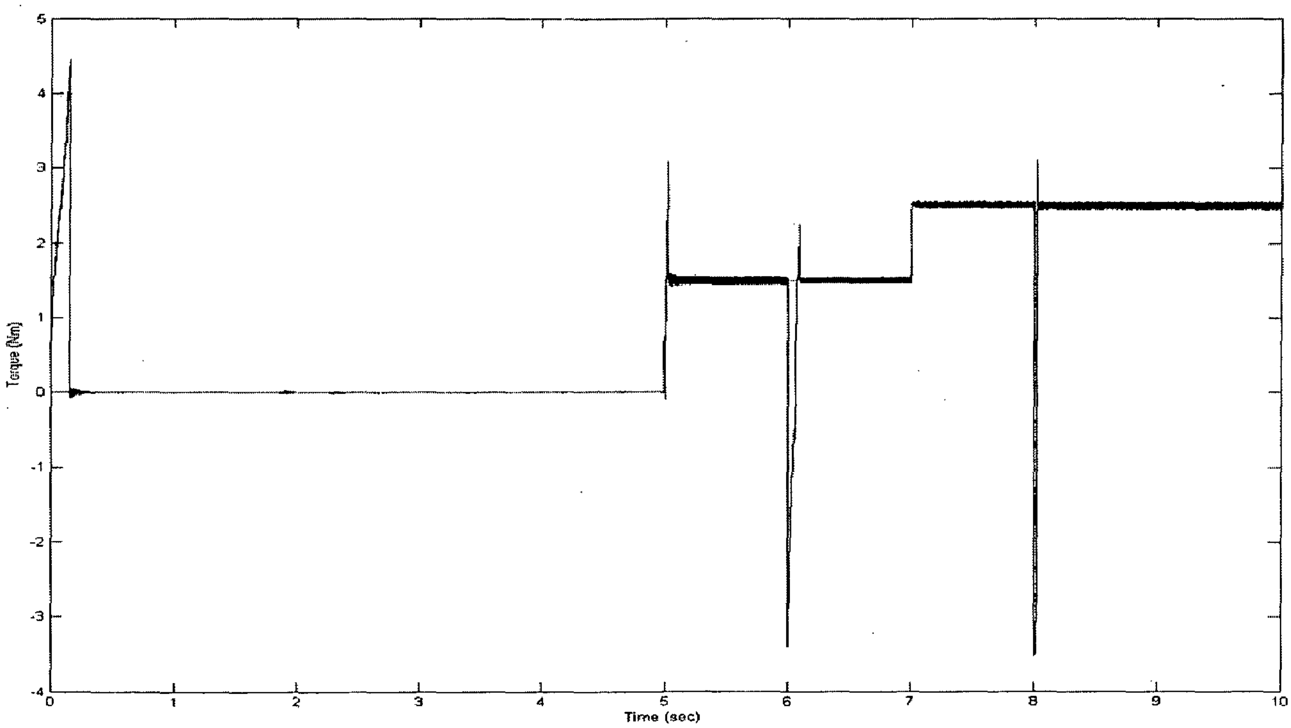

Fig.5.6 Torque response with Fuzzy controller

Fig 5.5 and Fig5.6 shows the speed and torque response of Induction motor drive using fuzzy controllers in which the reference speed is set by the step source externally. The speed of the drive was reached to the reference speed at different loads .By comparing the results with the results obtained by using PI controller the time taken by the drive while reaching the reference speed has been more. And the torque ripples are more incase of fuzzy controlled induction motor drive.

5.1.2 Performance of the 5.4HP (4KW) Vector controlled Induction motor drive

The simulation results of the 5.4HP Vector controlled Induction motor drive are presented in the figures from 5.7 to 5.12 for the operation of the drive for different speeds and loads by using Proportional and Integral (PI)controller and Fuzzy logic controller. Fig.5.7gives the speed response of induction motor drive ffor different loads set by the step source externally. The starting speed reference is 1400rpm.The speed of the drive reaches the reference speed with in 0.1sec and after that speed reference is changed from 1400 rpm to 1000 rpm. From the Fig. it is observed that speed reaches its reference speed within 0.1 sec..After 8 sec the speed set to negative speed and the speed settles in quick time.

The electromagnetic torque produced is shown in Fig 5.8. It settles the load torque which changing from noload to rated load . The response of the drive is fast. One can observe the sudden variations in the torque while changing the speed from one value to the other.Fig.5.9 shows the three phase winding currents and they are approximately sinusoidal and one can observe the variation of the frequency of the currents of the drive speedchanges.

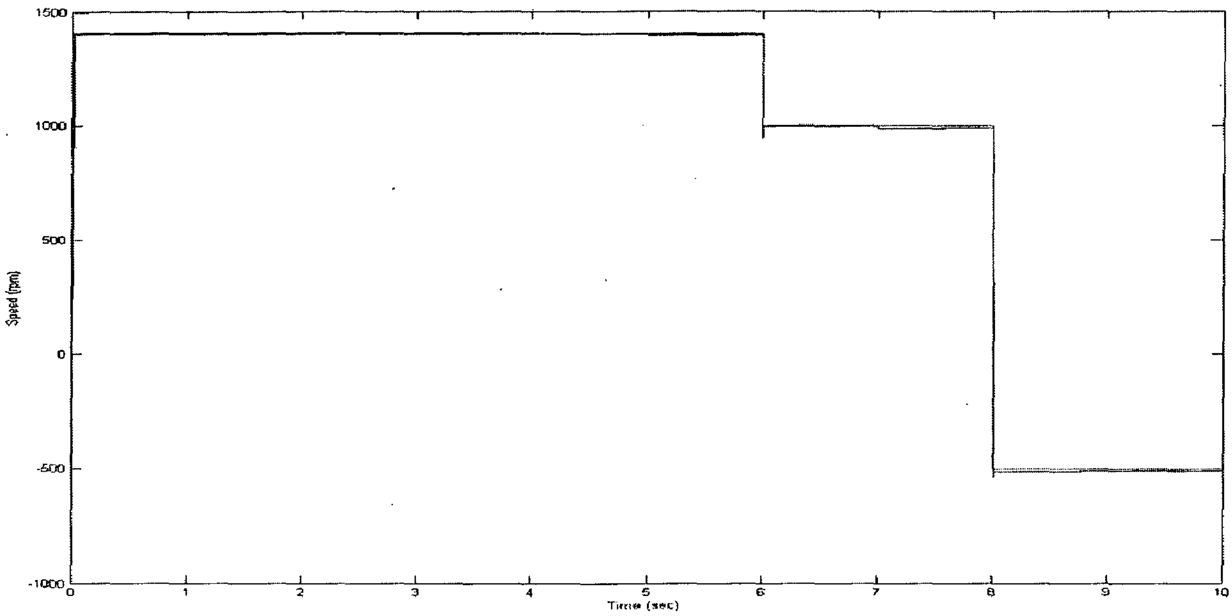

Fig.5.7 Speed response with PI controller

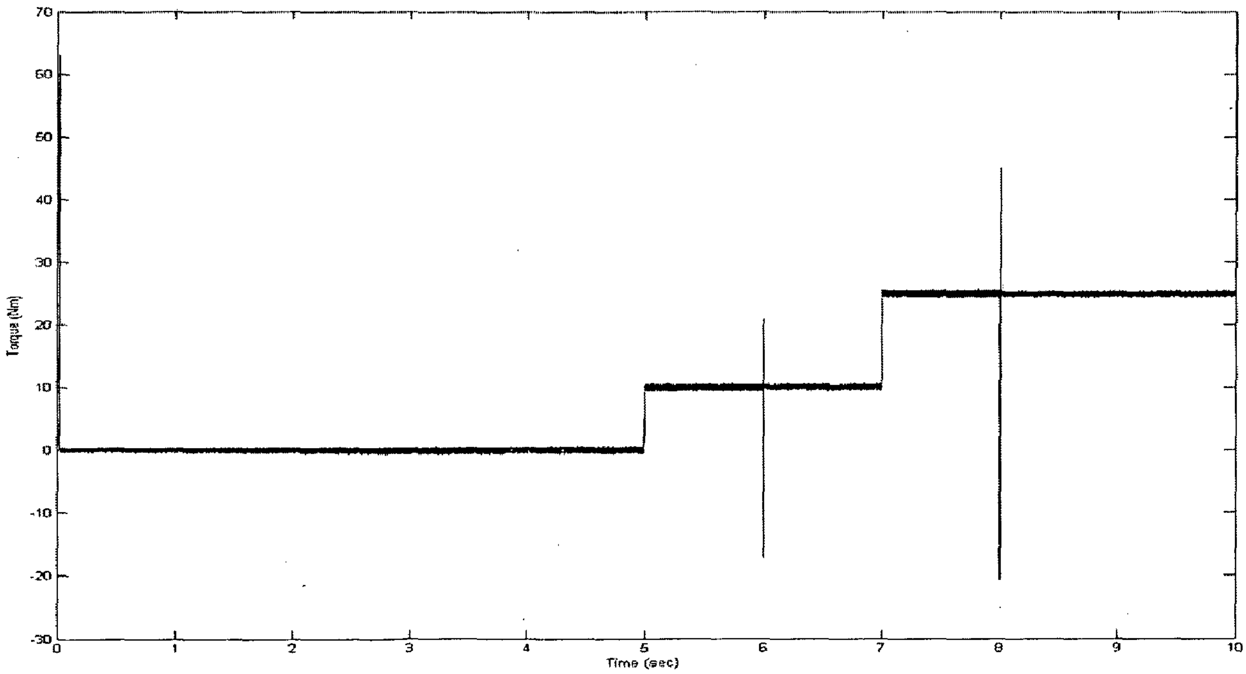

Fig.5.8 Torque response with PI controller

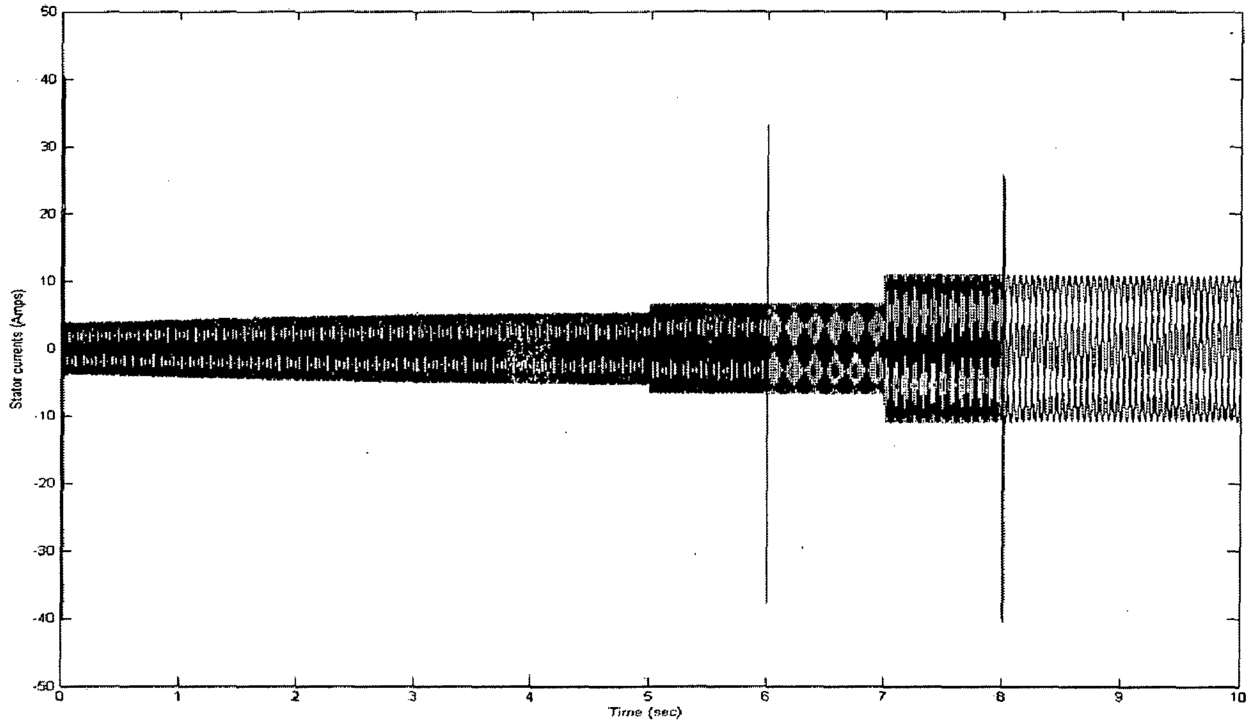

Fig.5.9 Stator currents response with PI Controller

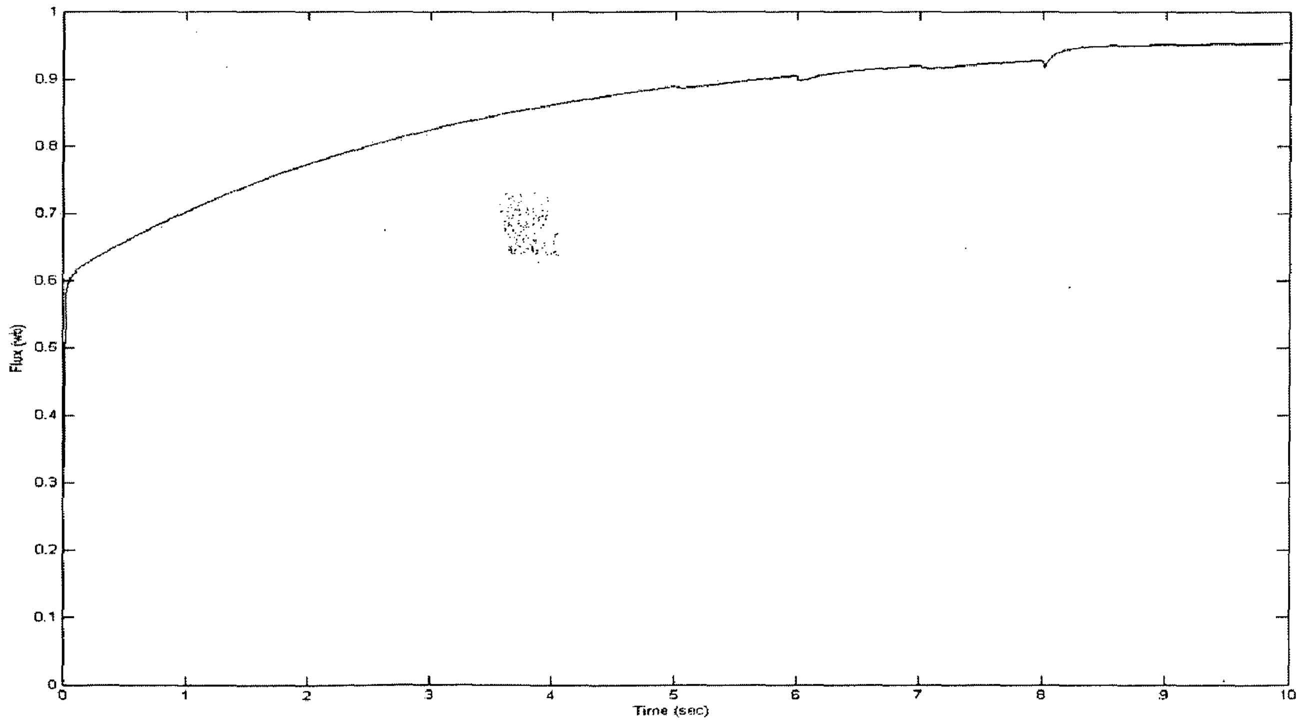

Fig.5.10.Flux response

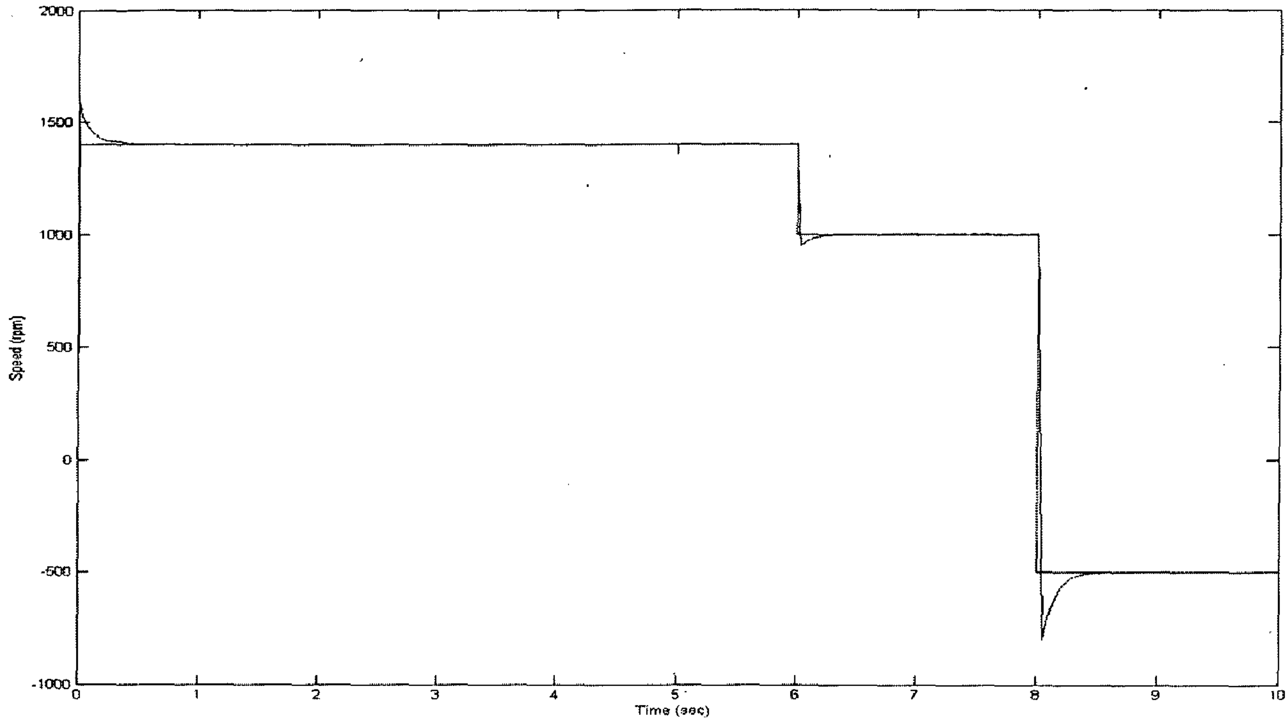

Fig.5.11 Speed response with Fuzzy Controller

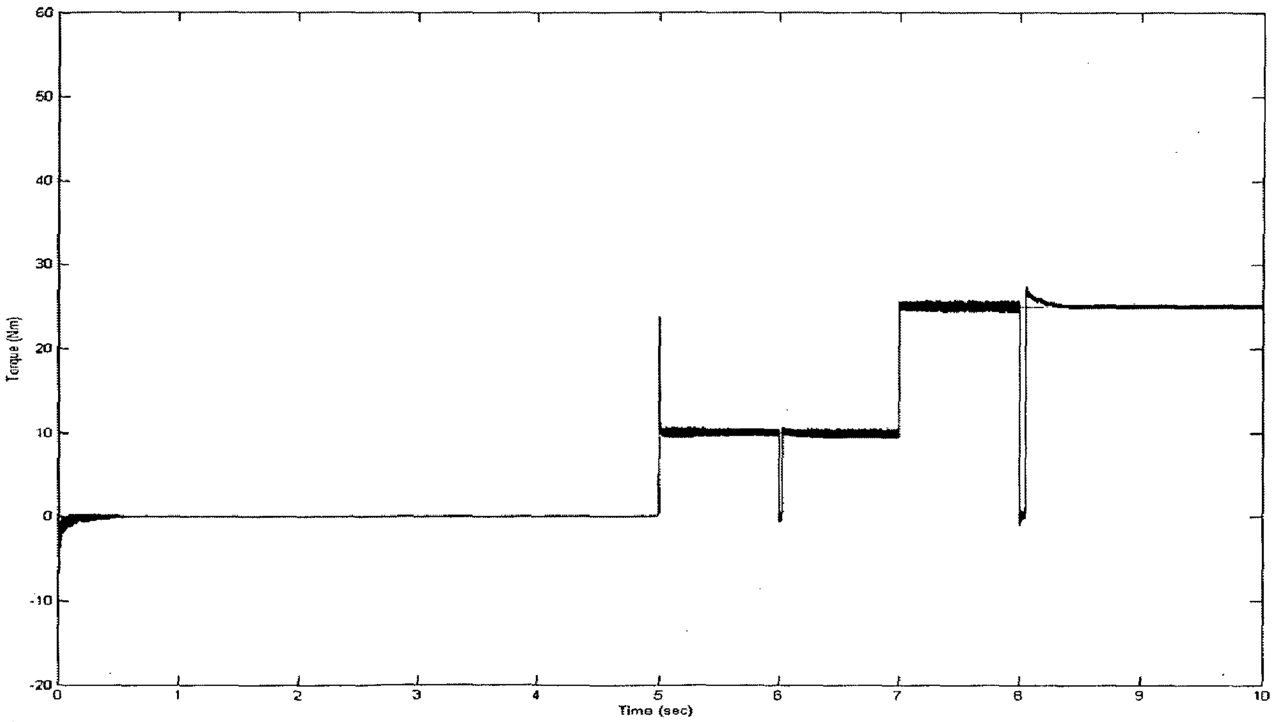

Fig.5.12 Torque response with Fuzzy Controller

5.2 OBSERVER BASED SENSOR LESS VECTOR CONTROL OF INDUCTION MOTOR DRIVE

The complete observer based speed sensor less induction motor drive is shown in Fig.4.3. The machine specifications are given in appendix-B. The speed response of the drive is of utmost significance and without sensing it directly; it is estimated and fed back for control action. By giving an input speed reference, the speed developed by the rotor is observed but without sensing the speed it is controlled i.e. sensor less operation. The criteria must be fast and smooth dynamic response. Both speed controls at no load, at load and speed reversal are discussed and also for a wide range of speeds i.e. from low speed region to high speed region and these are compared with the response without observer.

5.2.1 Response under No load Condition:

The simulation results of the 5.4HP Vector controlled Induction motor drive are presented in thefollowing figures for the operation of the drive for different speeds and loads by using Proportional and Integral (PI)controller and Fuzzy logic controller.

Fig.5.13gives the speed response of induction motor drive for different loads set by the step source externally. The starting speed reference is 1400rpm.The speed of the drive reaches the reference speed with in 5.5sec and after that speed reaches the steady state. Fig.5.16 represents the speed comparision of estimated speed,actual speed and reference speed. From that it is observed that estimted speed reached the reference speed within less time.

The electromagnetic torque produced is shown in Fig 5.14. The drive is operated at no load . From the figure it is observed that at no load operation the ripples produced in the electromagnetic torque are more. Fig.5.15 shows the three phase winding currents and they are approximately sinusoidal and one can observe the variation of the frequency of the currents of the drive speed changes .

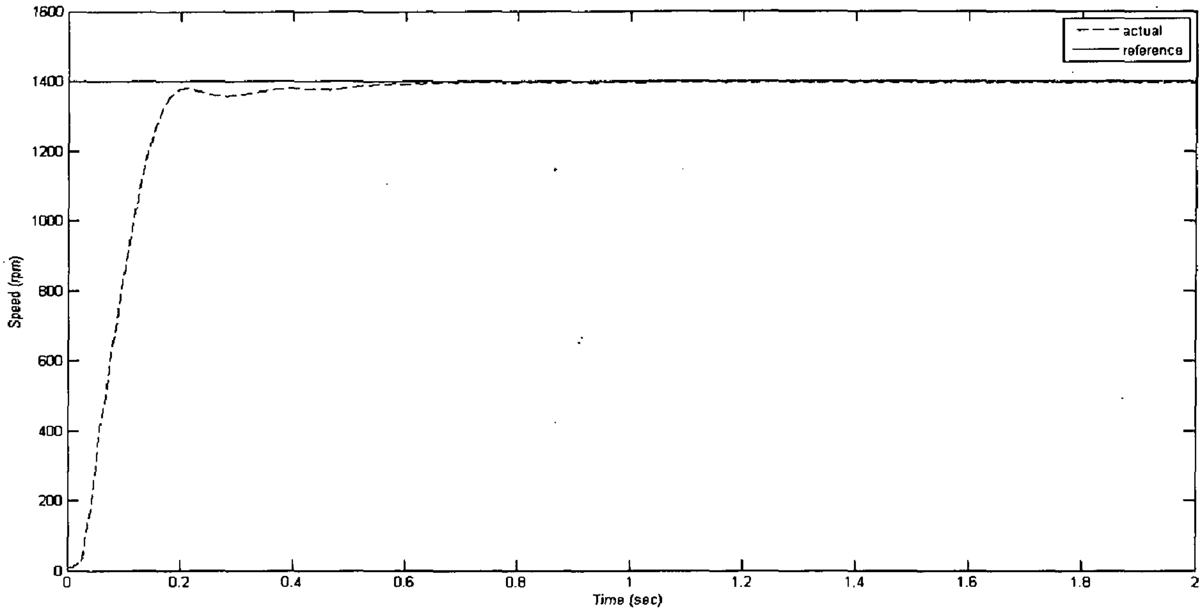

Fig.5.13. Speed response with PI controller

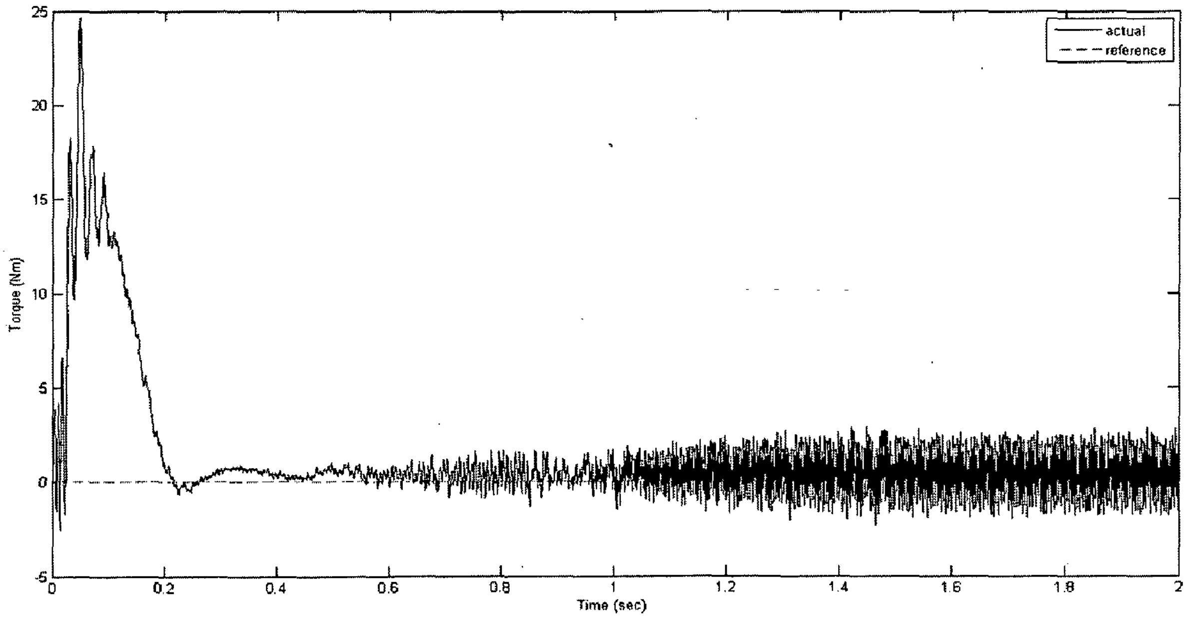

Fig.5.14. Torque response with PI controller

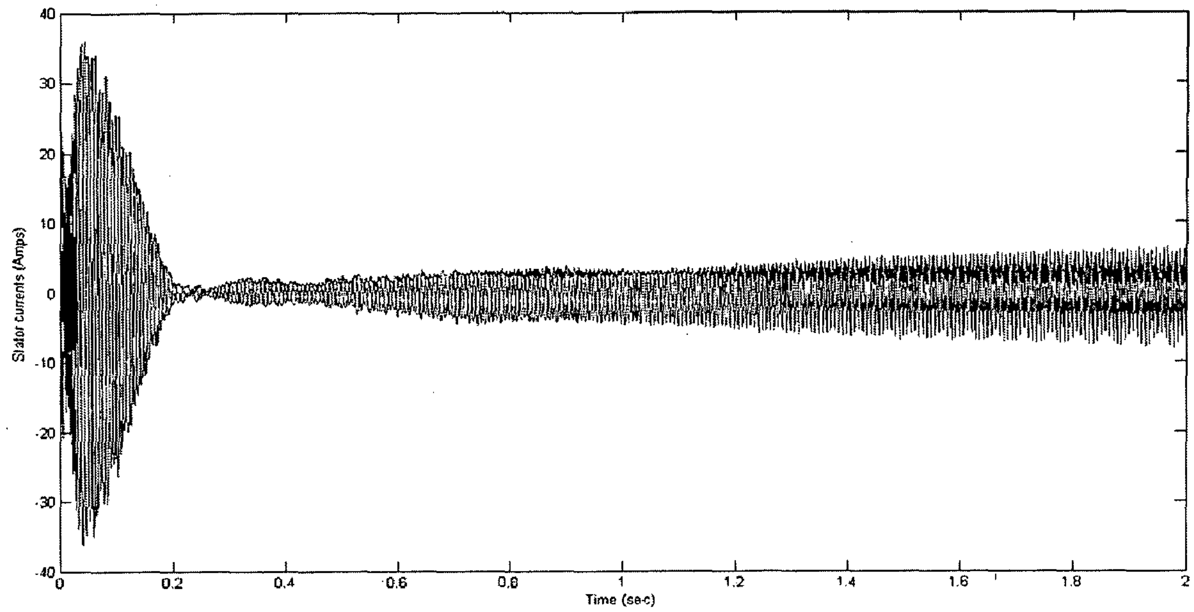

Fig.5.15 Stator currents response with PI Controller

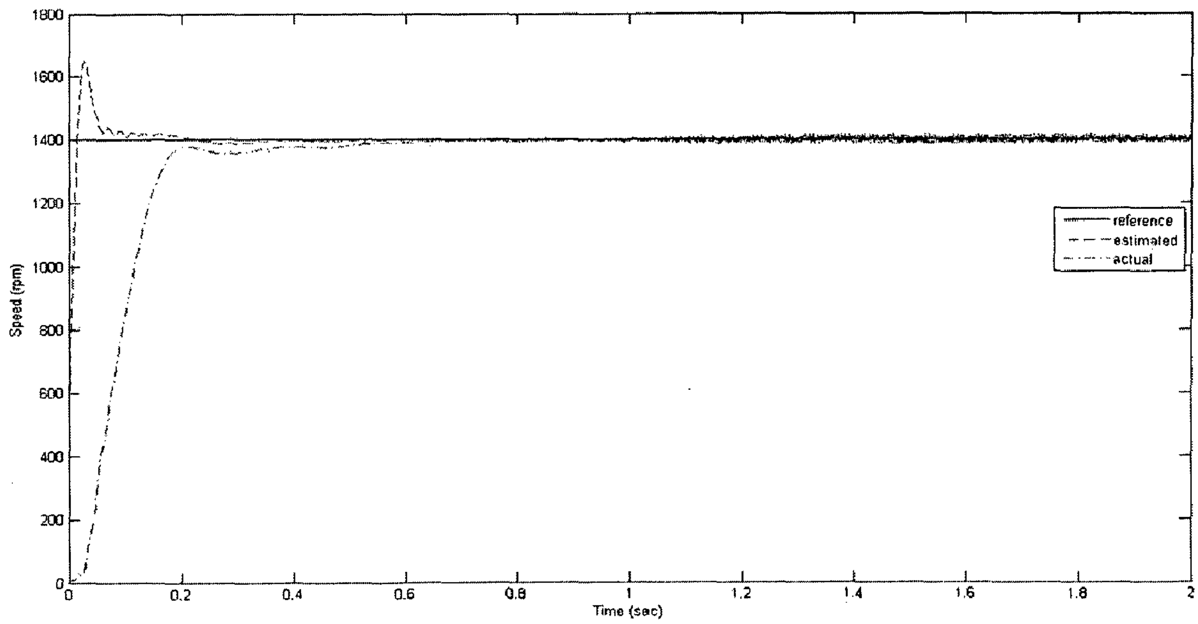

Fig.5.16 Speed comparison with PI Controller

The simulation results of the 5.4HP Vector controlled Induction motor drive are presented in the following figures for the operation of the drive for different speeds and loads by using Fuzzy logic controller. Fig.5.17 gives the speed response of the induction motor drive with starting speed set at 1400 rpm by using step source externally.by comparing the speed response using Fuzzy controller with speed response using PI controller,the speed reached the reference speed with in less time as compared to the later. And the peak overshoot occurred in case of drive using Fuzzy controller. Fig 5.20 gives the response of speed comparision of induction motor drive consisting of actual speed,estimated speed, and reference speed. One can observe that both the estimated and actual speed follows the reference speed at 0.3sec.

The electromagnetic torque produced is shown in Fig 5.18. The drive is operated at no load . From the Fig.it is observed that the starting torque produced is more in this case and the ripples reduced because of using Fuzzy controller. Fig.5.19 shows the three phase winding currents and they are approximately sinusoidal and one can observe the variation of the frequency of the currents of the drive speed changes .

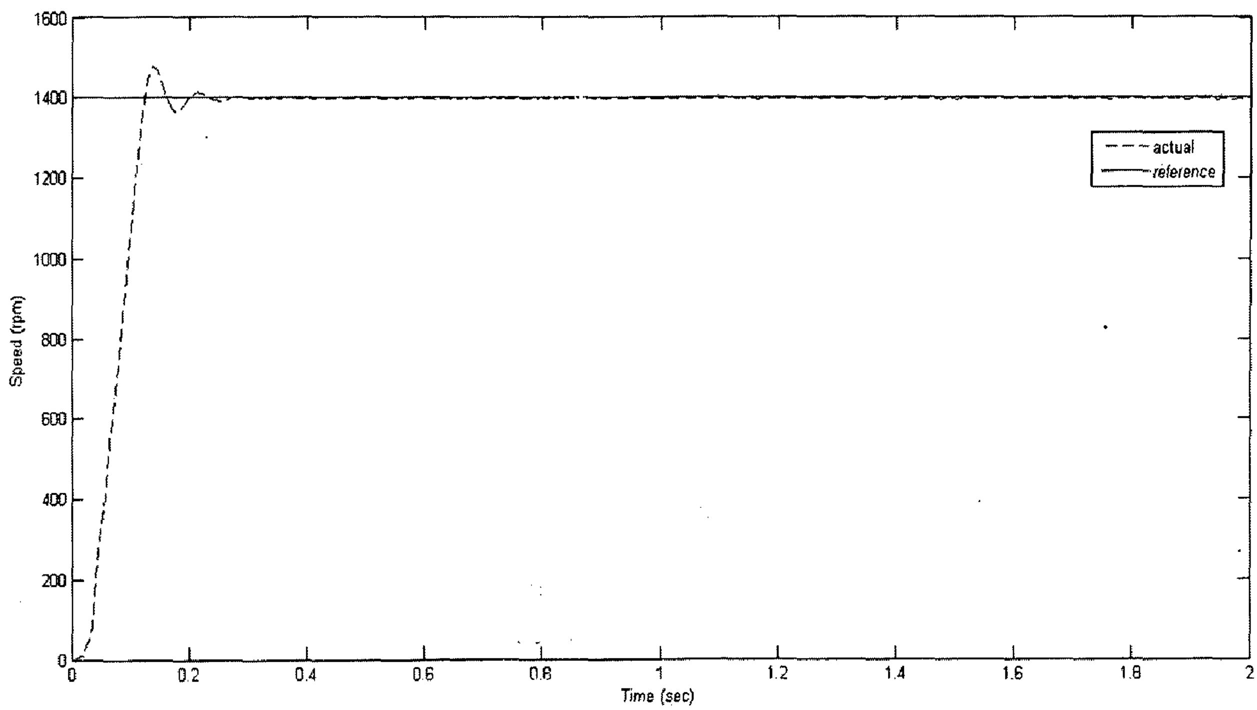

Fig.5.17 Speed response with Fuzzy Controller

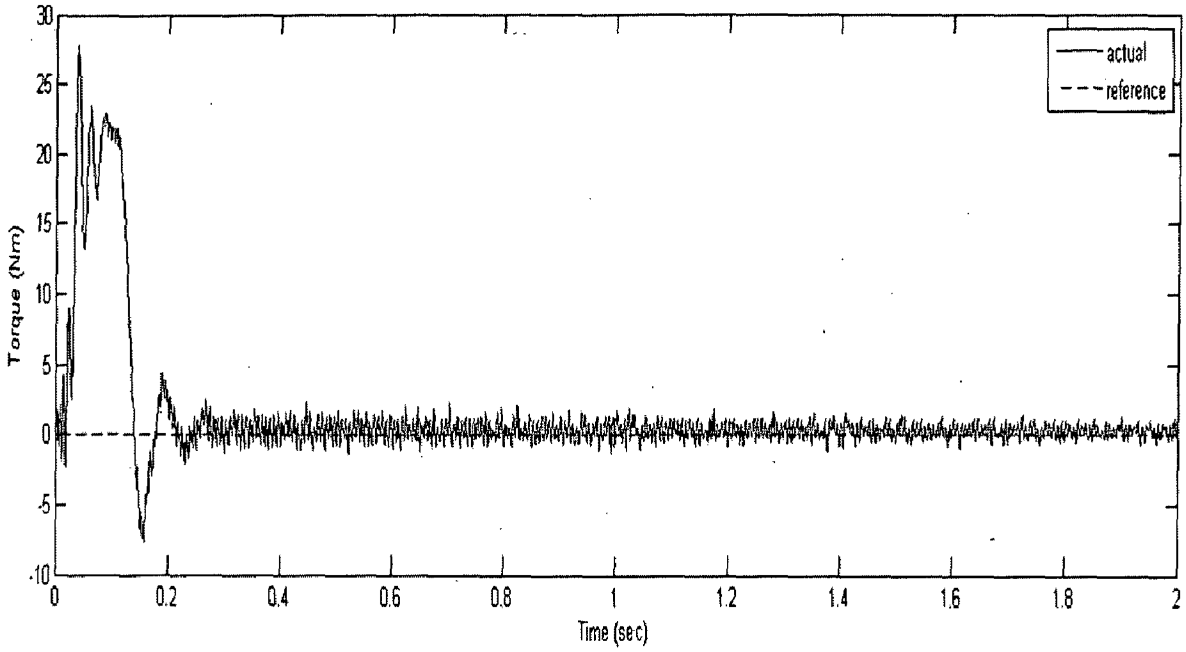

Fig.5.18 Torque response with Fuzzy Controller

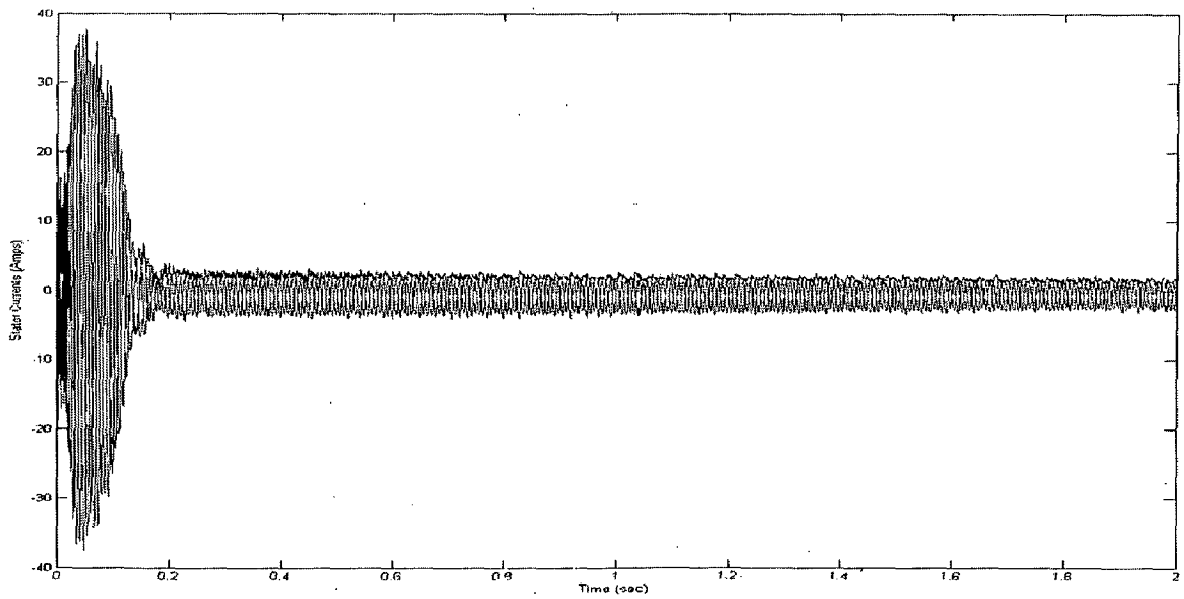

Fig.5.19 Stator currents response with Fuzzy Controller

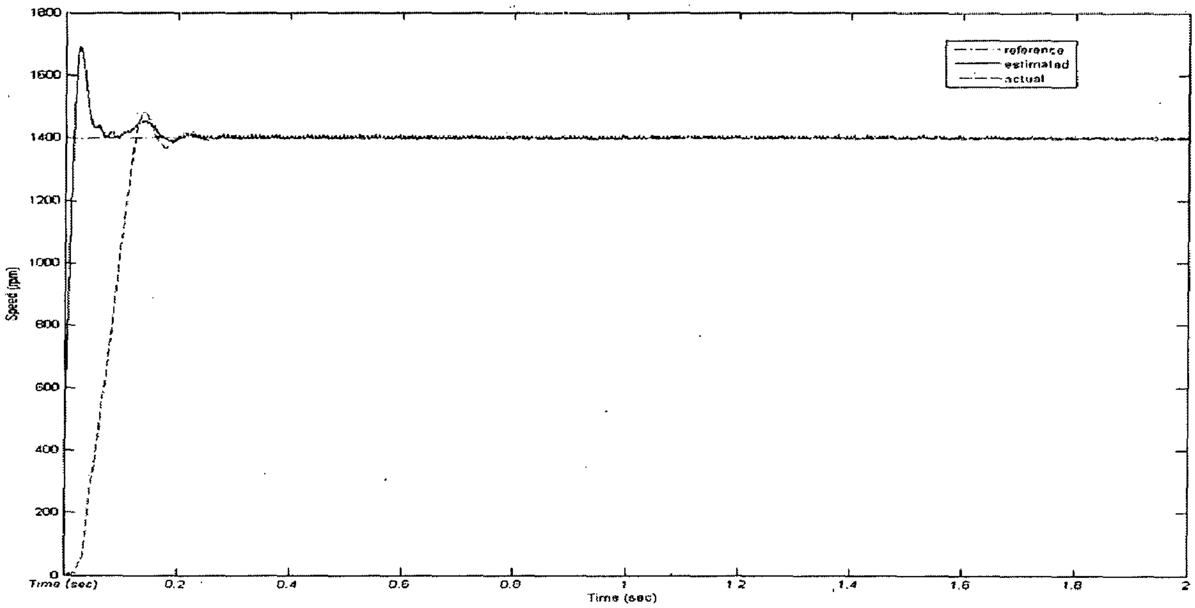

Fig.5.20 Speed comparison with Fuzzy Controller

The following figures gives the speed, torque and stator current responses of induction motor drive using Fuzzy controller for step change in speed at no load operation. For the simulation conducted, the speed input fed into the system remains at 1400 rpm for initial 1.2 sec and then goes to 800 rpm for remaining 0.8 sec. This is a step speed command. In the Fig.5.21, curve shows the actual speed developed by the motor and the other is the reference speed signal that is set by the observer. It shows that actual speed i.e. speeds of rotor is closely following the reference speed signal in the steady state. Fig. 5.24 shows the comparison of different speeds consisting of actual and estimated. It is observed that both estimated and actual speeds reaches the steady state value at 0.35 sec and for 800 rpm reached the reference value at 1.6 sec. It has taken more time to reach the steady state.

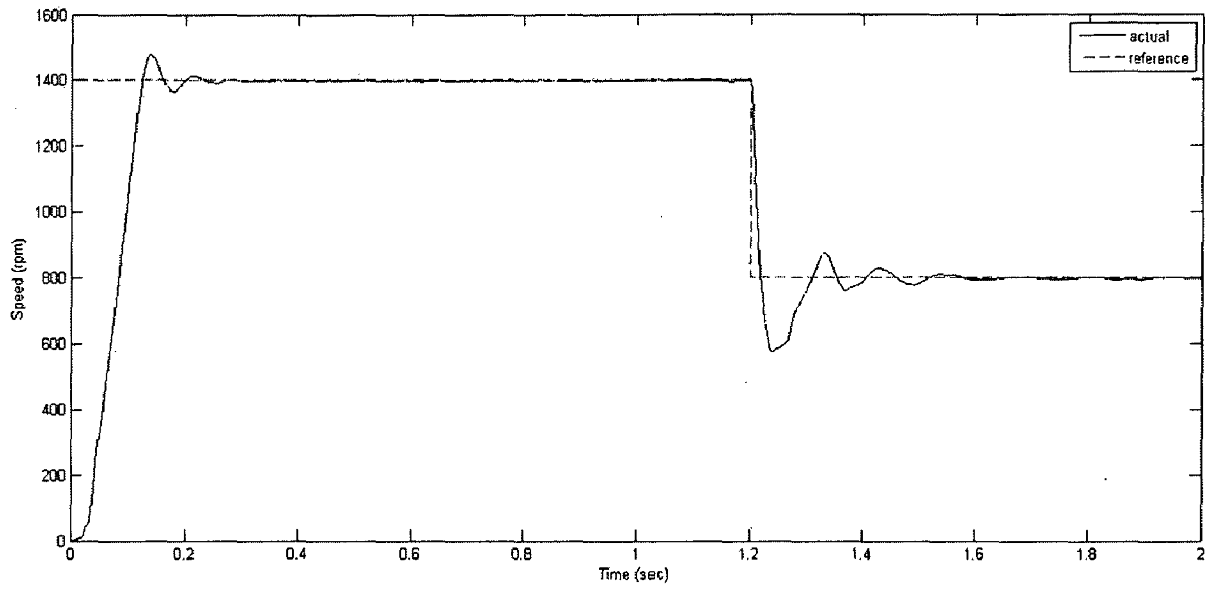

Fig.5.21 Speed response with Fuzzy Controller

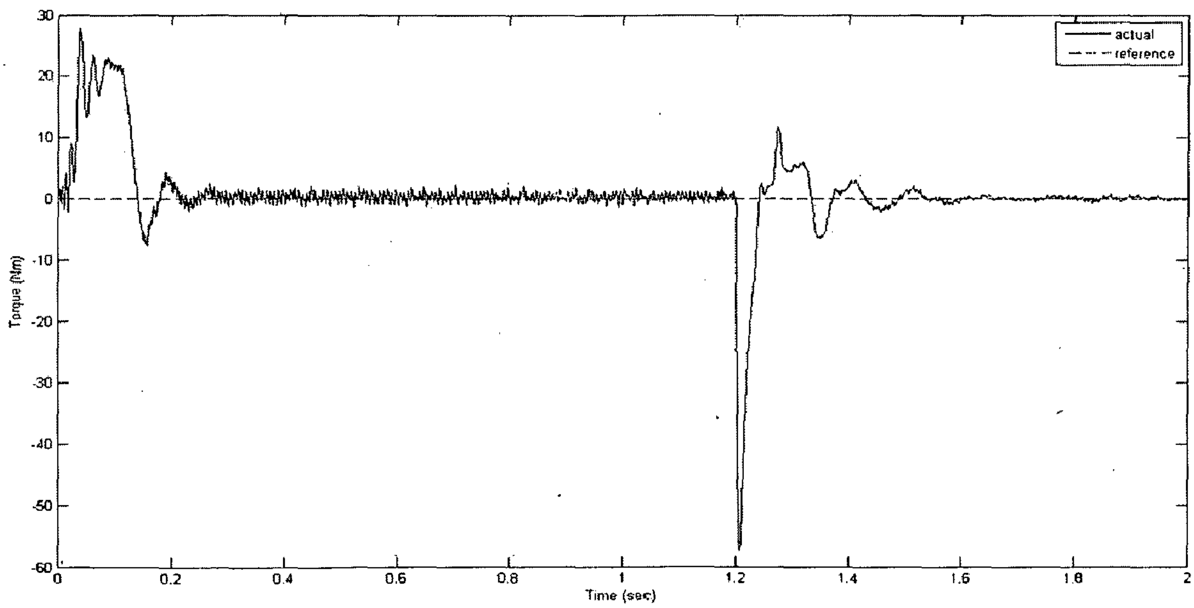

Fig.5.22 Torque response with Fuzzy Controller

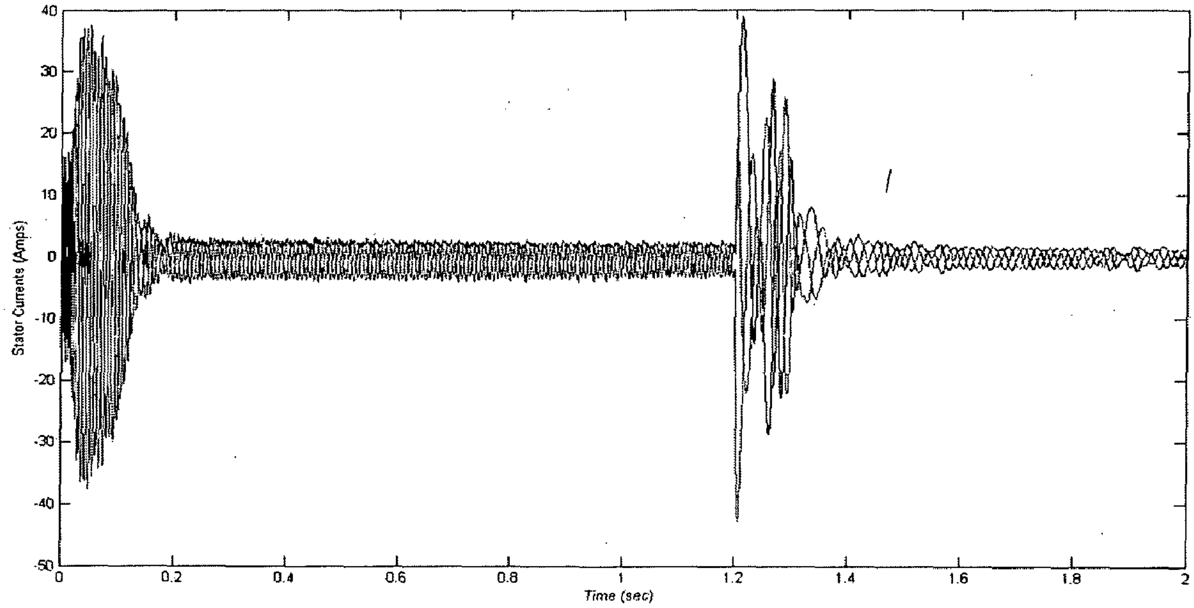

Fig.5.23 Stator currents response with Fuzzy Controller

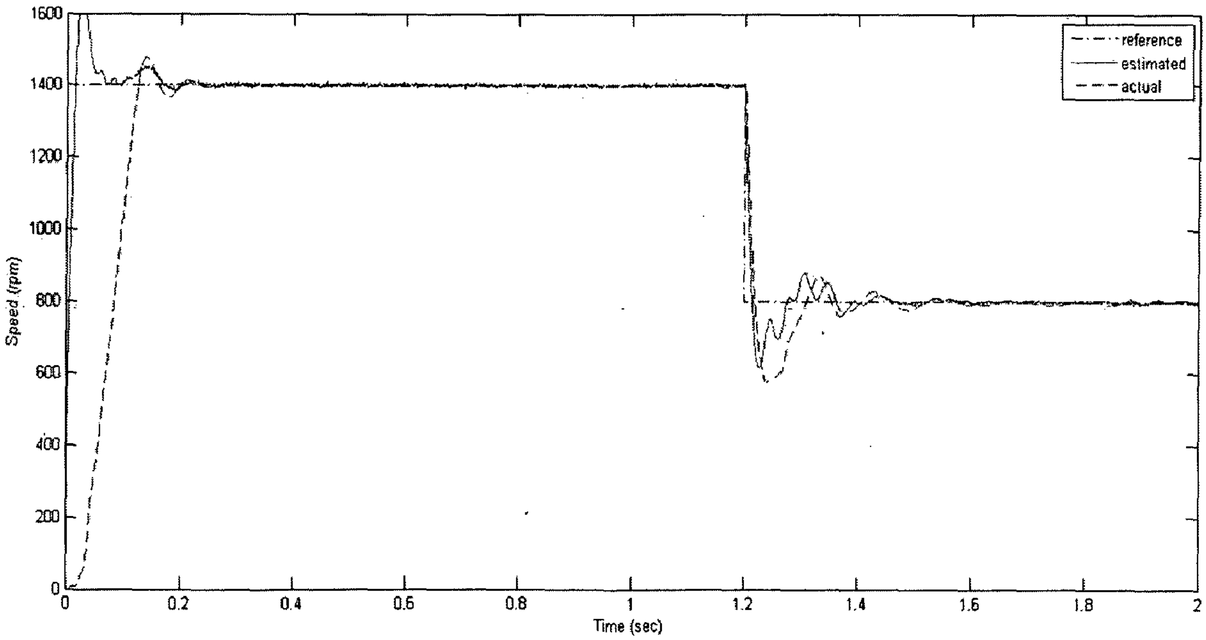

Fig.5.24 Speed comparison with Fuzzy Controller

The speed response for the different speed settings has been given below. Fig.5.25 gives the speed response of the induction motor drive with reference speed set at 1200 rpm. The drive is operated at no load. And the speed reaches the reference value at 5.5 sec., the ripples occurred in the response while it reaches the steady state are less. Fig 5.26 gives the comparison of estimated speed and actual speed. Initially the error between the estimated speed and actual speed is more and it is gradually reduced while speed reaches to the reference value. Fig.5.27 and 5.28 gives the speed response and speed comparison of induction motor with reference speed set at 800 rpm under no-load operation.

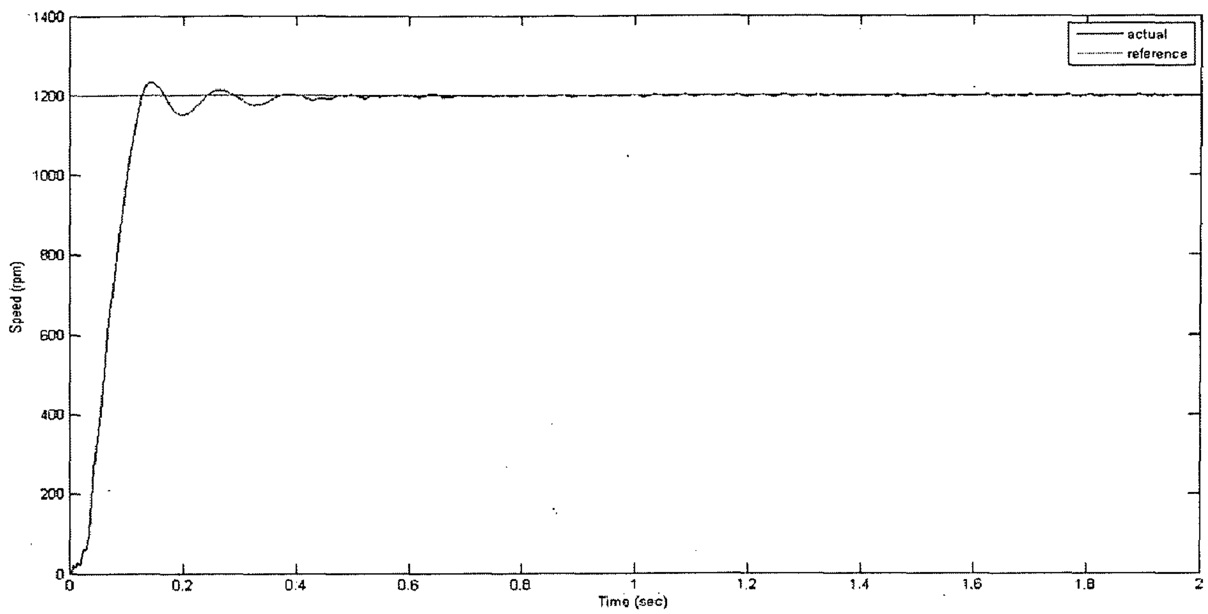

Fig.5.25 Speed response with PI Controller

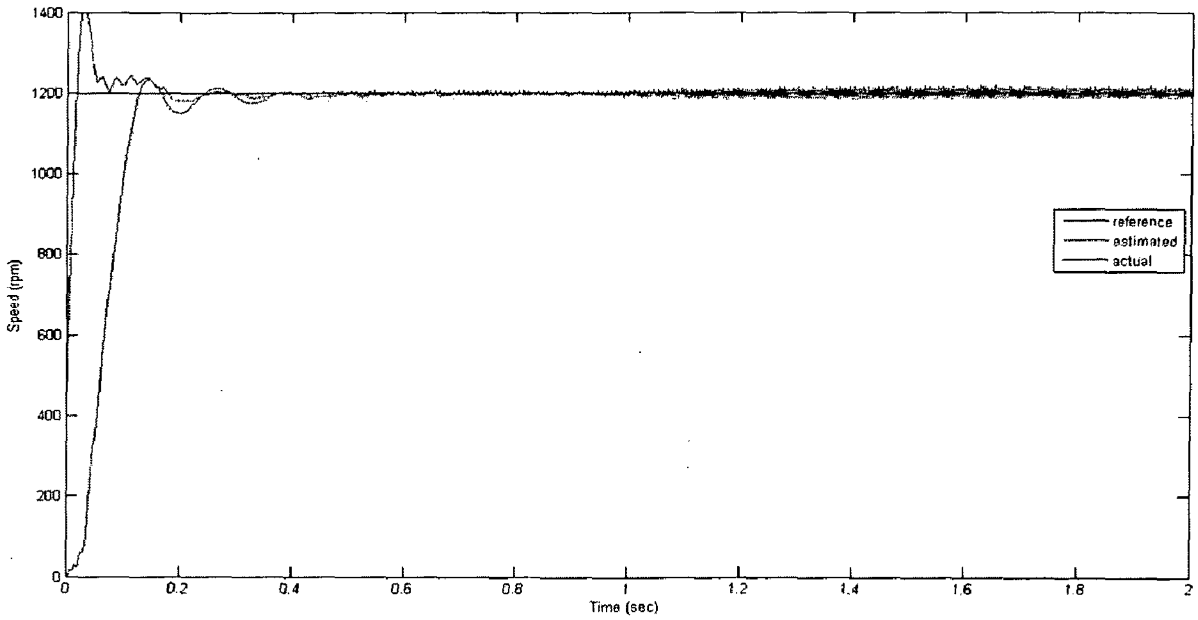

Fig.5.26 Speed comparison with PI Controller

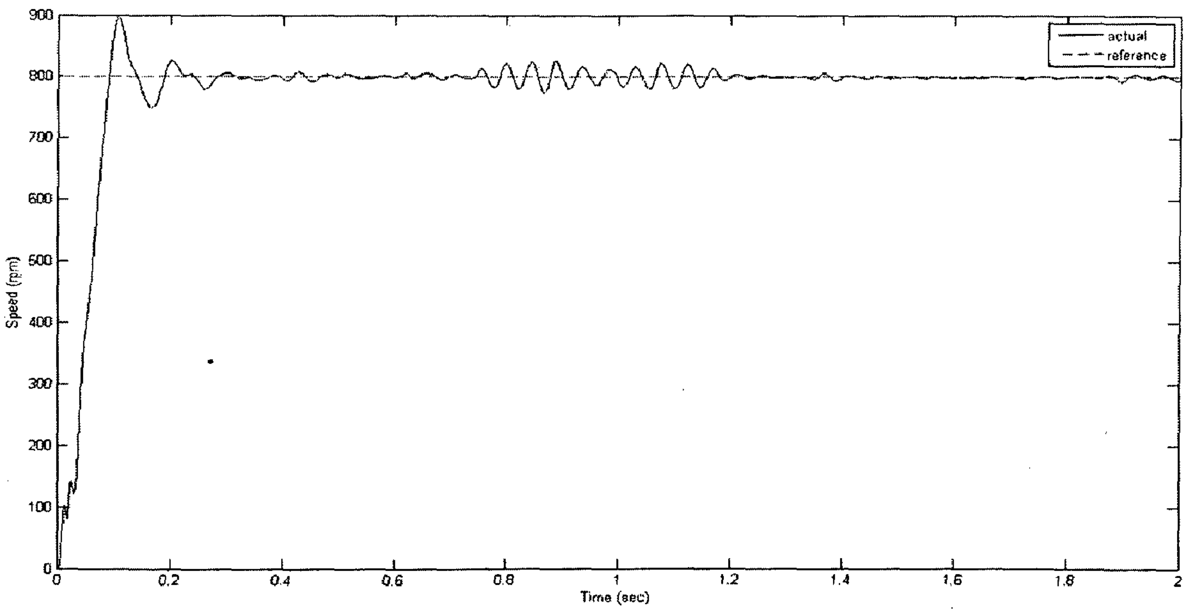

Fig.5.27 Speed response with PI Controller

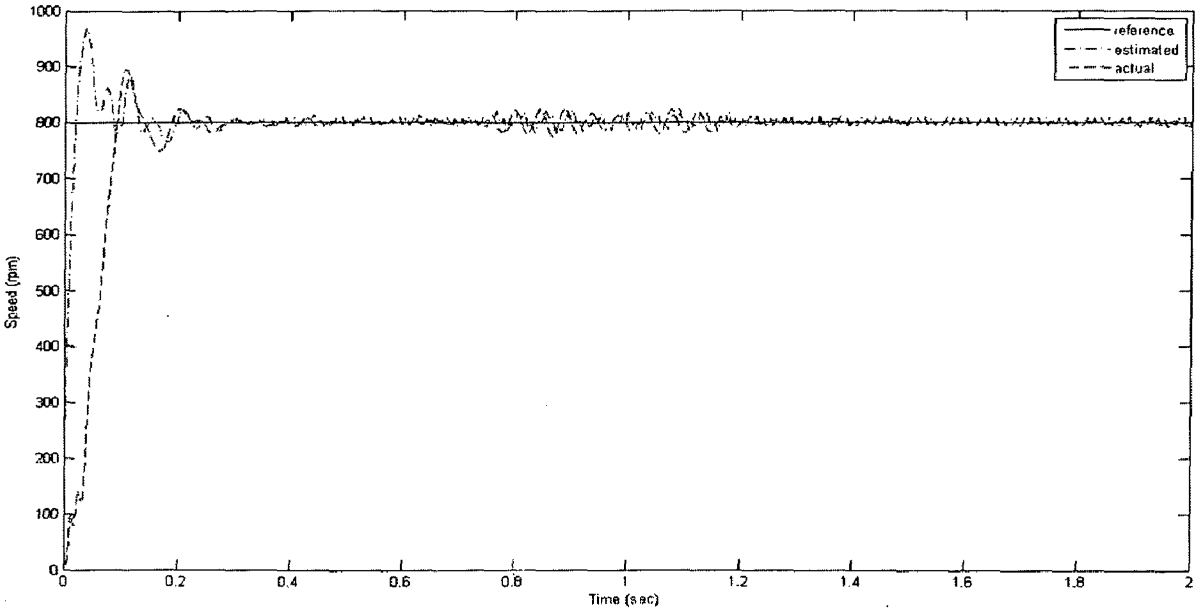

Fig.5.28 Speed comparison with PI Controller

5.2.2 Response under load condition:

For the analysis of torque response under load condition, a load torque of 10 Nm is applied at 1.2 sec and reference speed has changed from 1400rpm to 1000 rpm at 0.8 sec. The response of the system shows that under steady state condition developed torque follows the load torque, independent of speed command. If the speed reference changed to a new value, then also electromagnetic torque follows the load torque. Behavior of the motor under load condition is shown in the following figures. Fig.5.30 gives the comparison of actual speed and estimated speed under a load of 10 Nm applied on induction motor drive.

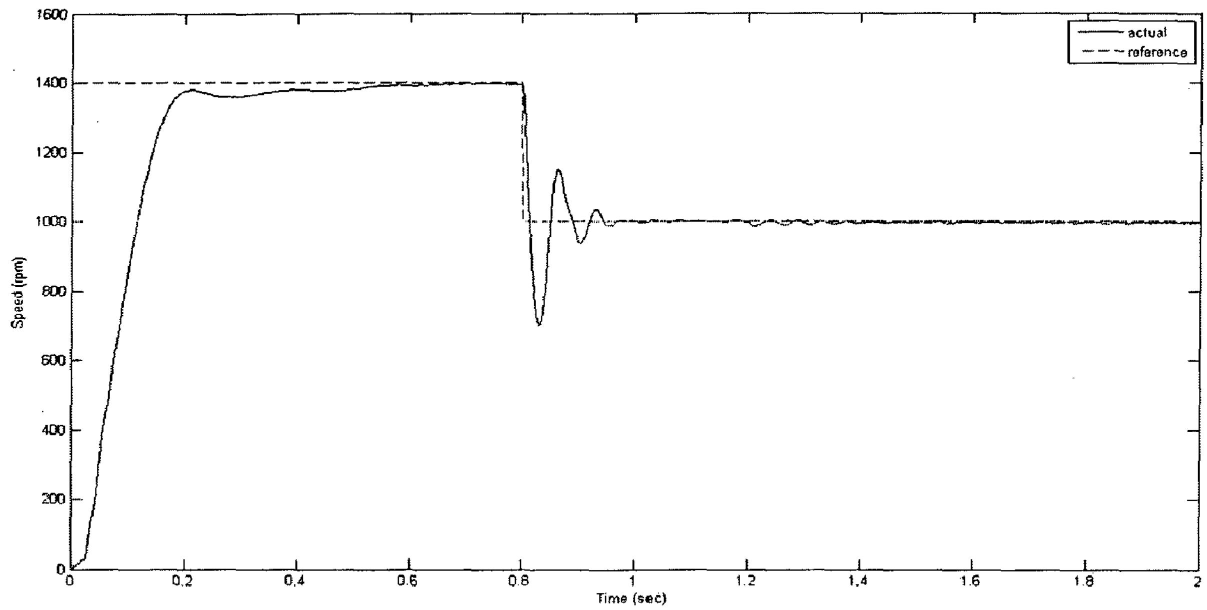

Fig.5.29 Speed response under load condition with Fuzzy controller

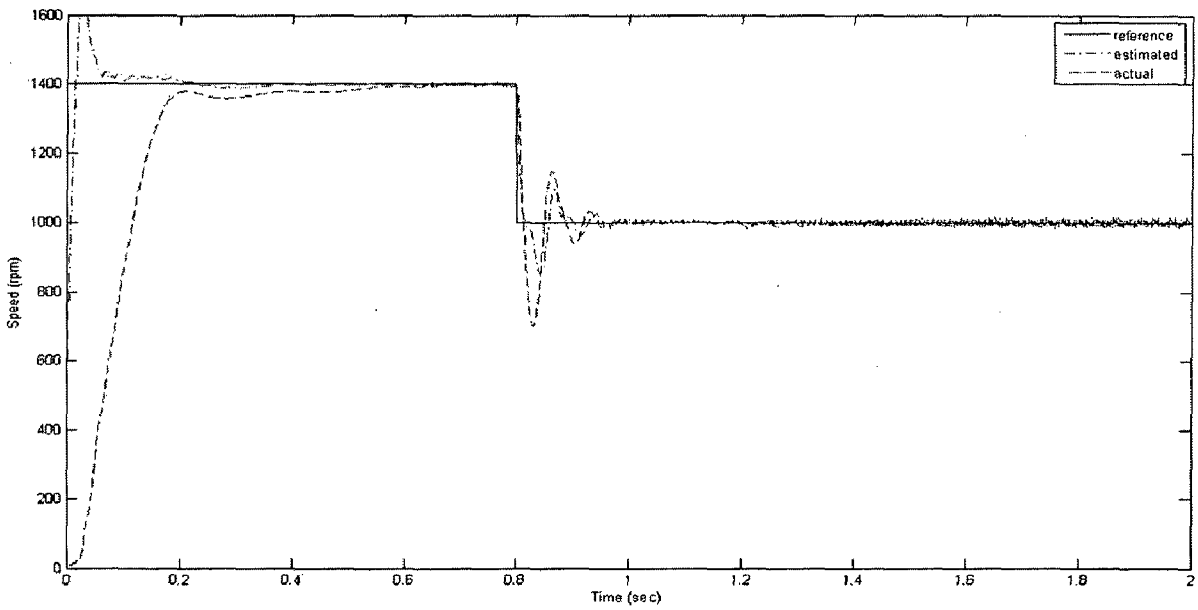

Fig.5.30 Speed comparison under load condition with Fuzzy controller

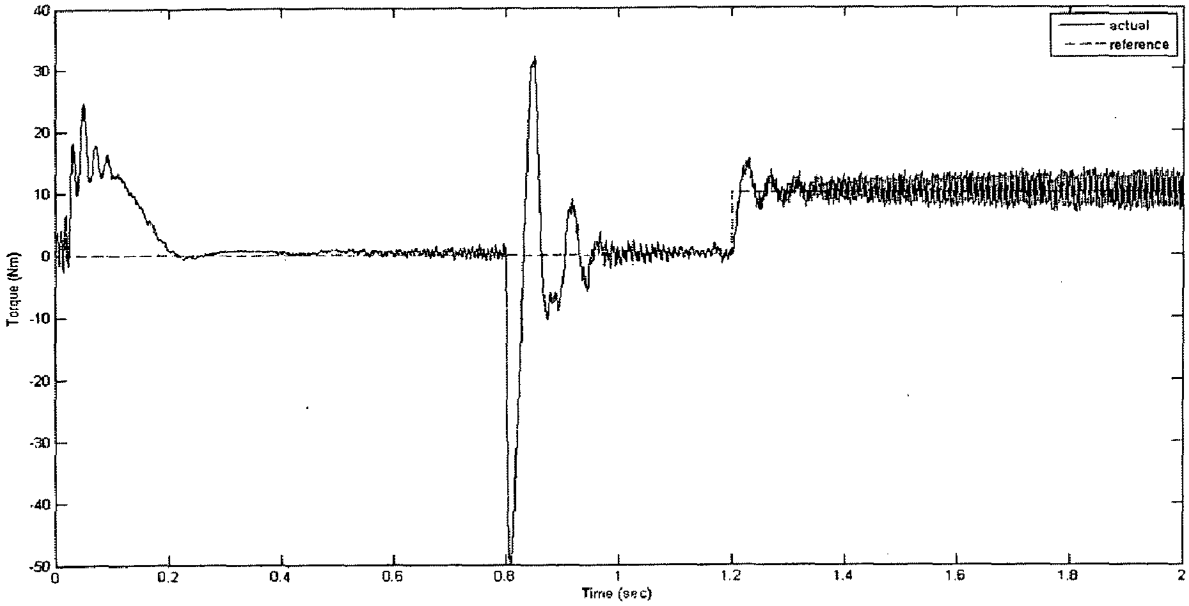

Fig.5.31 Torque response under load condition with Fuzzy controller

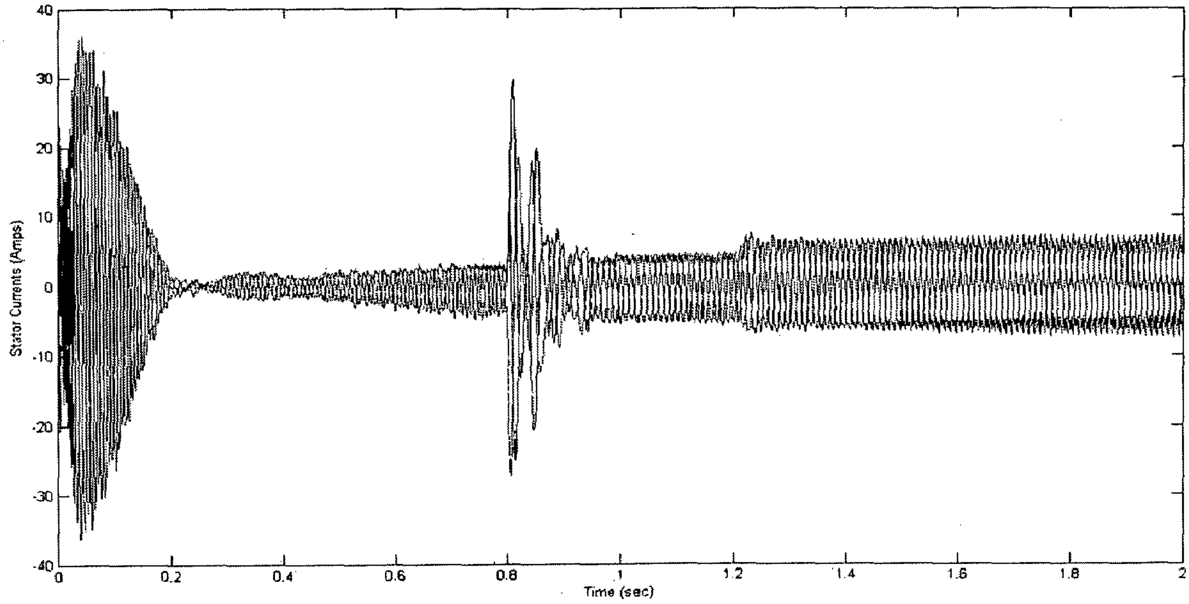

Fig.5.32 Stator currents response under load condition with Fuzzy controller

5.2.3 Response under Speed reversal operation

For simulation in reverse motoring operation, a step speed command of +1400 rpm to -600rpm is set as reference speed command. There is a step change in speed at 1.2sec. Load torque is set to zero. Speed response show in fig.5.33 shows that actual rotor speed closely follows the reference speed with minimum oscillations during positive reference speed i.e. during acceleration. But during deceleration the oscillations occurred. This response is obtained with PI controller .where as with Fuzzy controller the oscillation gets reduced and the actual rotor speed closely follows the reference speed with minimum oscillations under all conditions i.e. during acceleration, deceleration and steady state. Fig. 5.34 shows the comparison of actual speed and estimated speed

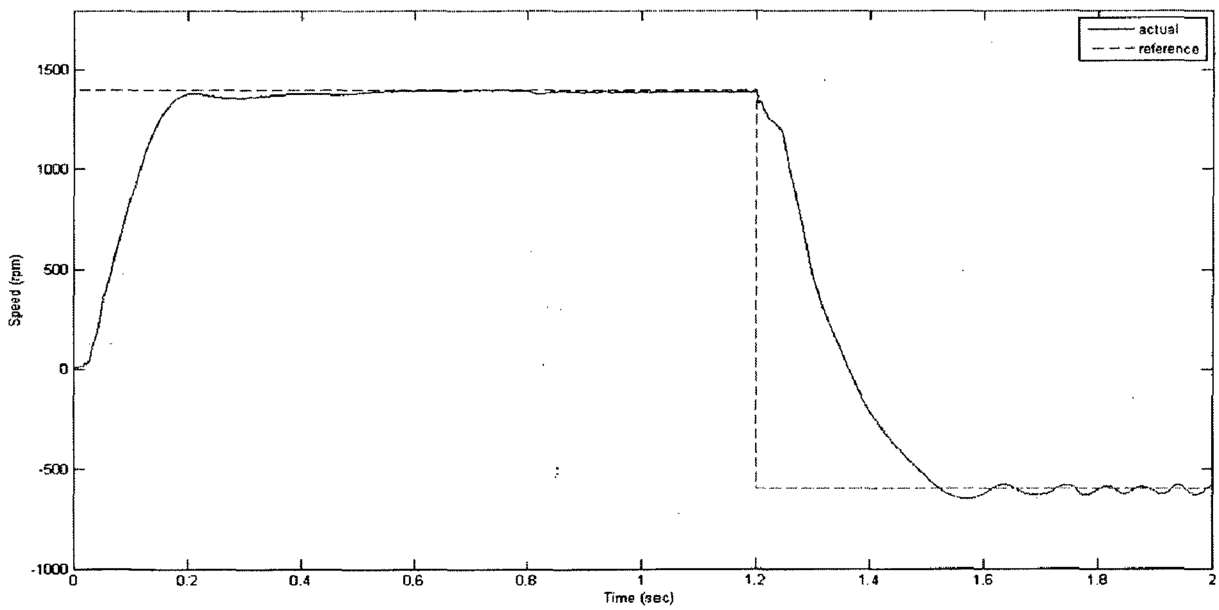

Fig.5.33. Speed reversal response with PI Controller

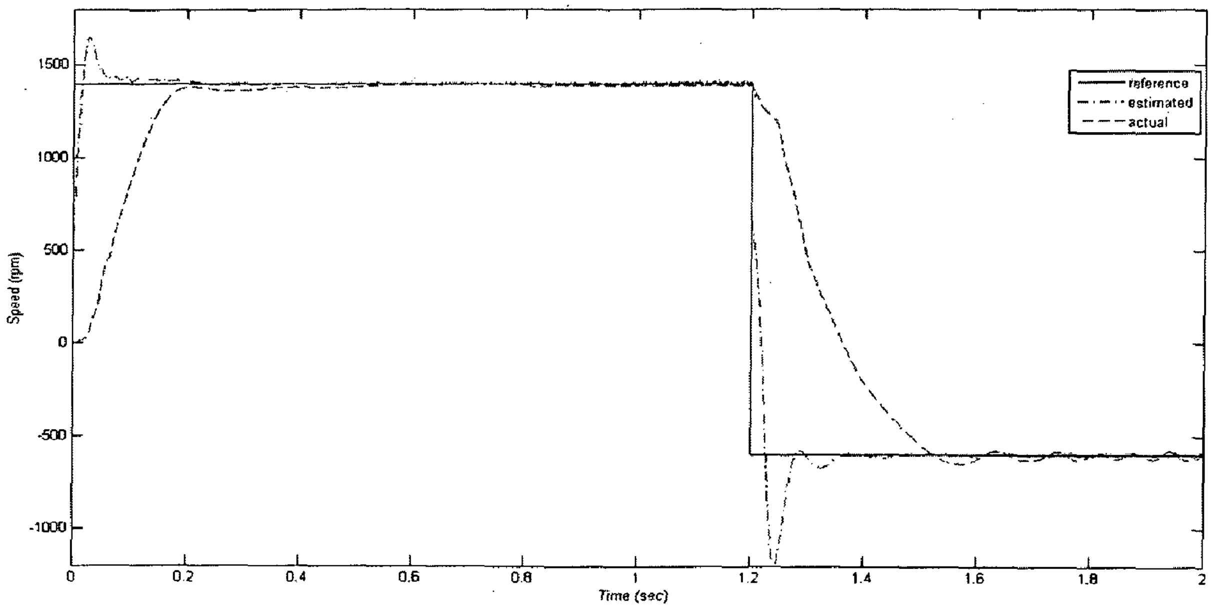

Fig.5.34. Speed comparison with PI Controller

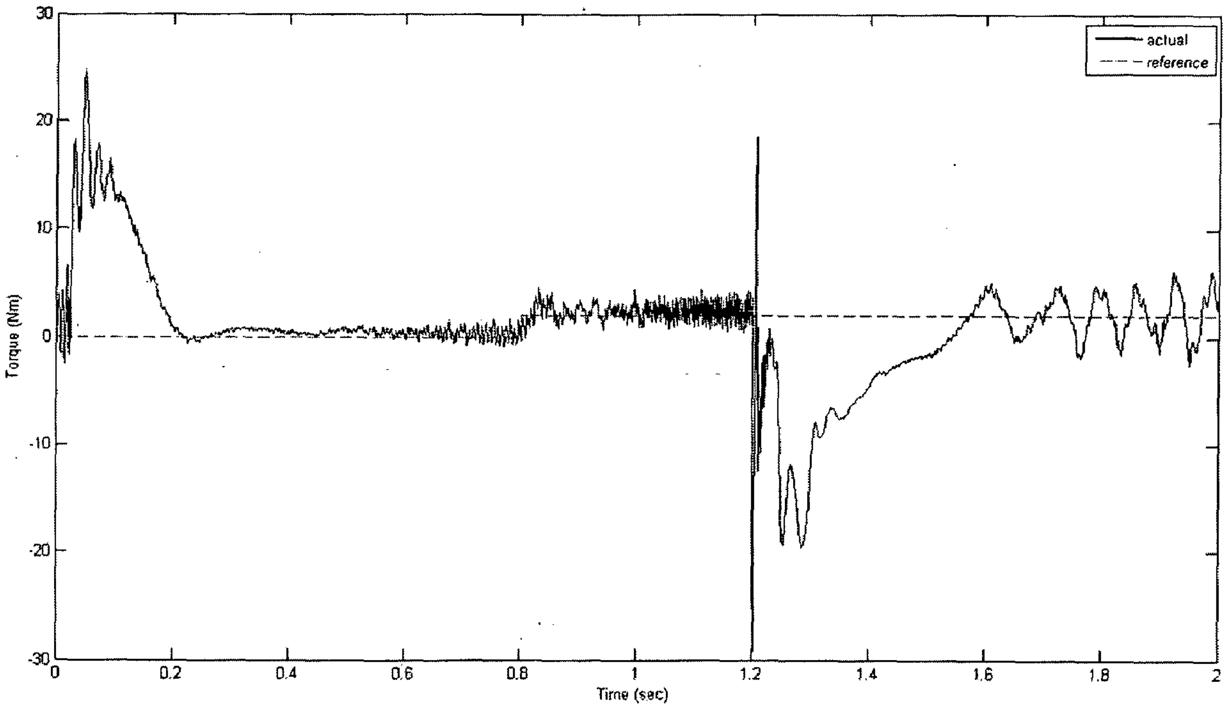

Fig.5.35 Torque response during Speed reversal with PI Controller

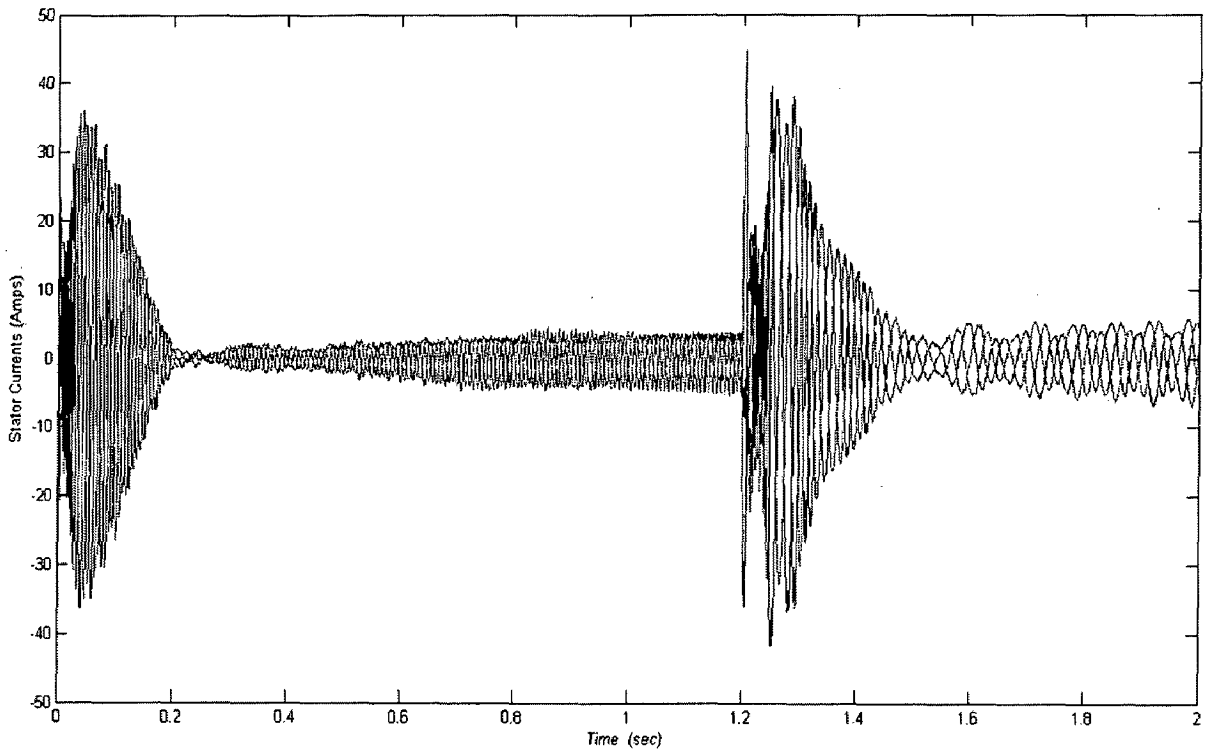

Fig.5.36. Stator currents response during speed reversal with PI Controller

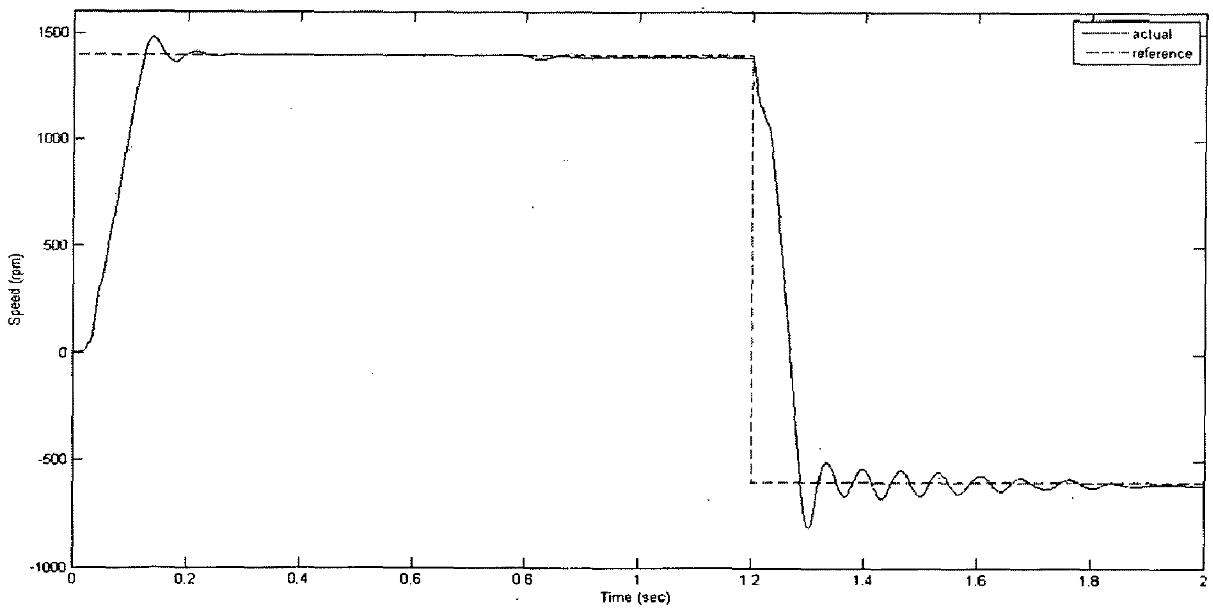

Fig.5.37. Speed reversal response with Fuzzy Controller

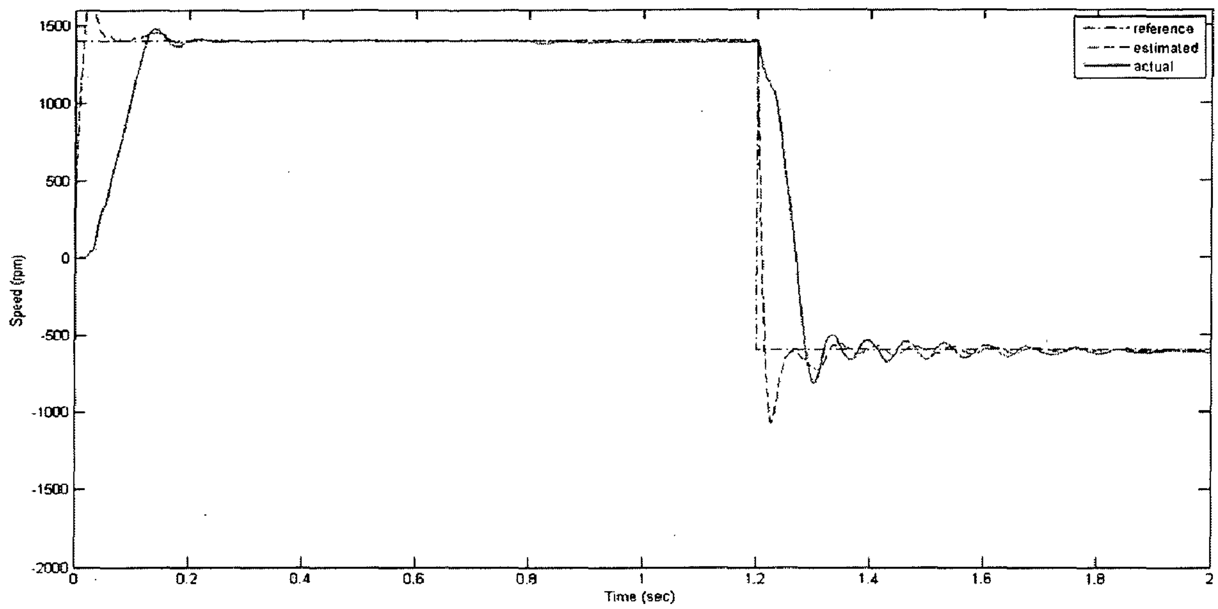

Fig.5.38. Speed comparison during reversal with Fuzzy Controller

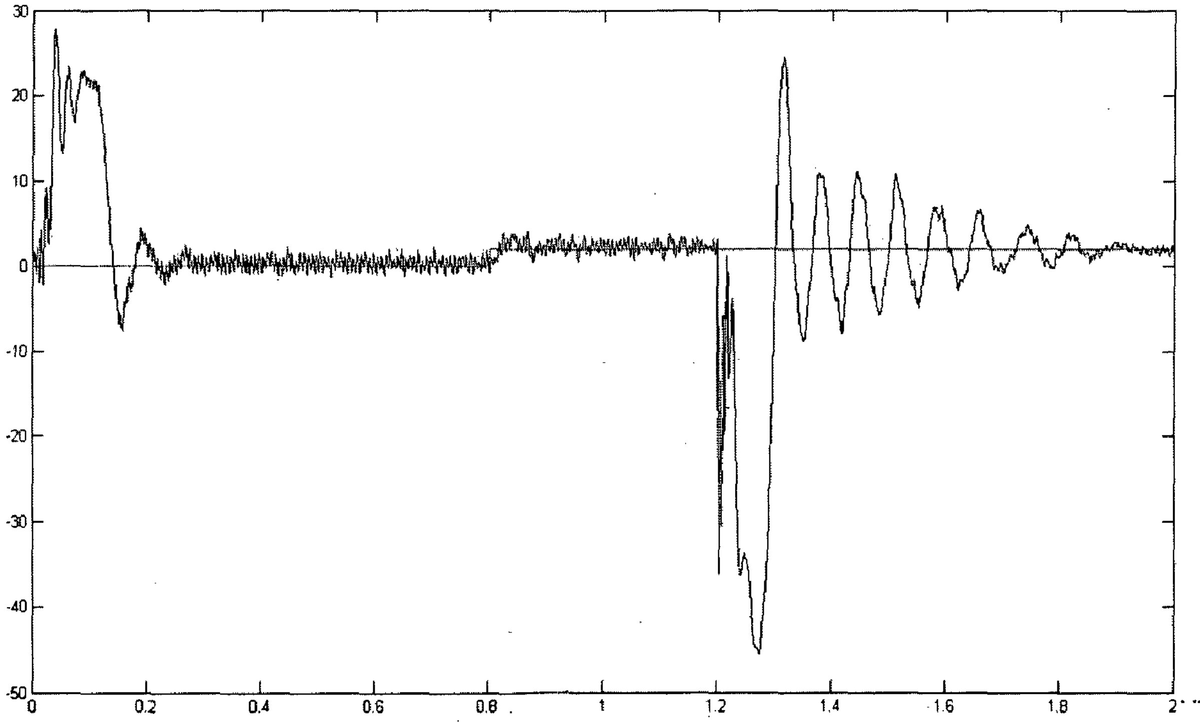

Fig.5.39. Torque response using Fuzzy controller

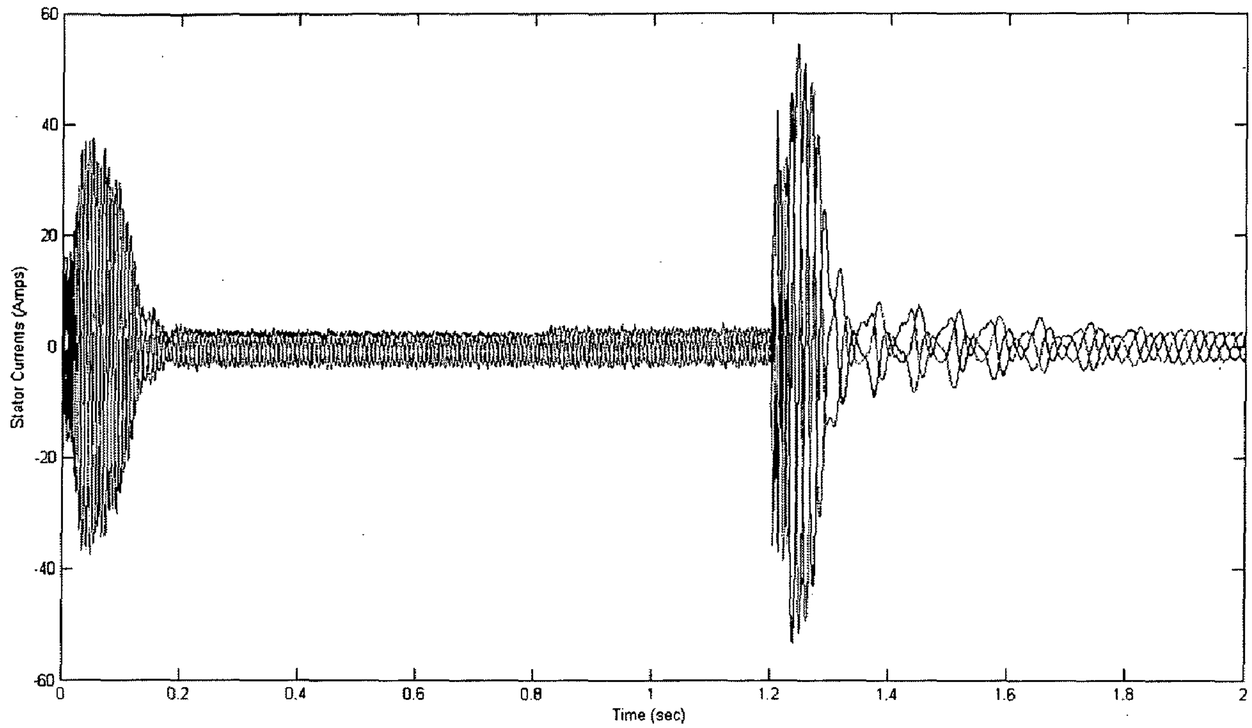

Fig.5.40. Stator currents response with Fuzzy Controller during speed reversal

CONCLUSIONS AND SCOPE FOR FURTHER WORK

In the present dissertation the performance analysis of an observer based sensor less speed controlled squirrel cage induction motor drive employing a controlled voltage source PWM inverter has been carried out. The study is based on the mathematical model of vector controlled induction motor discussed in Chapter 2. Through the simulation study performed by using the MATLAB/SIMULINK package, it is established that the field oriented control structure in conjunction with a Fuzzy controller, provides a faster dynamic response in view of settling time and peak overshoot. Further, the speed response doesn't depend upon direct measurement of rotor speed but it is estimated by the observer and good performance is achieved. The variation of the magnitude and frequency of the stator current of the motor, in desired manner, results in quicker accelerating torque. If required, this may also lead to regenerative action, as well as change in the phase sequence. This happens in response to disturbances in the drive structure such as perturbations in one or more of the system variables like the load on the shaft, the reference speed setting. With the above improved dynamic response of the induction motor drive system it well suited for a number of applications involving variable speed such as the process industries, machine tools, textiles industries, paper mills, lifts and traction.

Writing dedicated software, using for example C or C++ to investigate the performance of an induction motor, is time consuming. In this dissertation a commercially available software package (MATLAB/SIMULINK) has been used to study the performance of an induction motor driven by observer based sensor less speed IM drive. The speed, flux and torque PI controllers have successfully been implemented.

Following conclusions are derived from the results obtained.

- Good Tracking of reference speed is possible without sensing the speed directly than using sensor.
- Decoupled control of direct axis component and quadrature axis component of the stator current is possible.
- Speed response is quite fast.
- Improved speed response than indirect vector control without observer.

SCOPE FOR FURTHER WORK

The speed estimation is adversely affected by stator resistance variations due to temperature and frequency changes. This is particularly significant at very low speeds where the calculated flux deviates from its set values. Therefore, it is necessary to compensate for the parameter variation in sensor less induction motor drives, particularly at very low speeds. To improve this method of estimating both the shaft speed and stator resistance of an induction motor has to be done.

The speed control performance of the observer based speed sensor less controlled induction motor can be improved if the controller design includes intelligent control concepts. The intelligent controllers, such as *neuro-fuzzy controllers*, do not need any mathematical model of the plant. They are based on the plant operator experience and are very easy to implement.

Simulation results of one of the machines of the department have also been recorded. The same need to be validated through experimentation for the complete model. The above can be implemented using Digital controllers. The space-harmonics and time- harmonics components must be taken into account for more accurate transient analysis of the machine.

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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:

Stator resistance (Rs) = 11.124 Ω

Rotor resistance (Rr) = 8.9838 Ω

Stator leakage Inductance (Lls)= 33.36 mH

Rotor Leakage Inductance (Llr)= 33.36 mH

Mutual Inductance (Lm)= 490.45 mH

Moment of inertia (J)= 0.0018 J Kg-m^2

APPENDIX – B

MOTOR PARAMETERS-II

Type : Squirrel cage induction motor

Phase : 3

Power : 5.4 Hp

Voltage : 400 V

Poles : 4

Frequency: 50 Hz

Equivalent circuit parameters:

Stator resistance (Rs) = 1.405Ω

Rotor resistance (Rr) = 1.395 Ω

Stator leakage inductance (Lls) = 5.839 mH

Rotor leakage inductance (Llr) = 5.839 mH

Mutual inductance (Lm) = 172.2 mH

Moment of inertia (J) = 0.0131 Kg-m^2