Servo Control of Induction Motor Drive using Indirect Vector Control

Thesis submitted in partial fulfillment of the requirements for the award of degree of

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in
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Certificate

I hereby certify that the work which is being presented in the thesis entitled, **"Servo Control of Induction Motor Drive using Indirect Vector Control"**, in partial fulfillment of the requirements for the award of degree of Master of Engineering in *Power Systems & Electric Drives* submitted in Electrical & Instrumentation Engineering Department of Thapar University, Patiala, is an authentic record of my own work carried out under the supervision of Mr. Yogesh K. Chauhan.

The matter presented in this thesis has not been submitted for the award of any other degree of this or any other university.

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

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CONTENTS

ABSTRACT

The vector control of induction motor is used instead of conventional scalar control aims at decoupling the torque and flux producing components of the stator current under all operating speed and load conditions. Consequently, the drive can be tuned for quasi – instantaneous tracking of load and reference changes. The speed information is normally applied to a Fuzzy PI controller, thus generating the necessary current vector command for achieving independent torque control.

The d-q model of the induction motor is used in the stationary reference frame for the analysis purpose. The indirect field – oriented induction motor drive where the torque and flux producing components of the stator current are generated from a Fuzzy PI controller. The complete system is simulated in Sim Power Sim/Simulink (MATLAB). The proposed system track command speed and rejects the step disturbances in load torque with zero steady state error and very least transient response error which provides instantaneous torque control. The four quadrant operation is simulated through the developed controller, and the excellent results shows the superb working of fuzzy rule based controller of indirect vector control of induction motor.

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LIST OF SYMBOLS USED

a	120º operator.
$i_{ri}(t)$	Rotor current per phase.
$\bar{i}_r$	Space phasor of the rotor current expressed in the rotor reference frame.
$\bar{i}_r'$	Space phasor of the rotor current expressed in the stator reference frame.
$i_{si}(t)$	Stator current per phase.
$\bar{i}_s$	Space phasor of the stator current expressed in the stator reference frame.
$\bar{i}_s'$	Space phasor of the stator current expressed in the rotor reference frame.
L_m	Three phase magnetizing inductance.
L_r	Total three phase rotor inductance.
$\overline{L}_r$	Rotor self-inductance.
L_{r1}	Leakage rotor inductance.
L_{rm}	Rotor magnetising inductance.
L_s	Total three phase stator inductance.
$\overline{L}_s$	Stator self-inductance.
L_{sm}	Stator magnetising inductance.
L_{s1}	Leakage stator inductance.
$\overline{M}_r$	Mutual inductance between rotor windings.
$\overline{M}_s$	Mutual inductance between stator windings.
$\overline{M}_{sr}$	Maximal value of the stator- rotor mutual inductance.
p	Derivation operator.
P	Pair of poles.
R_r	Rotor Resistance.
R_s	Stator Resistance.
s	Slip.

$1/s$	Integration operator.
Te	Instantaneous value of the electromagnetic torque.
T_{pc}	Instant torque referred to the nominal torque and in percentage.
$Ts = Tz$	Sampling time.
$u_{ri}(t)$	Rotor voltage per phase.
$\bar{u}_r$	Space phasor of the rotor voltage expressed in the rotor reference frame.
$\bar{u}_r'$	Space phasor of the rotor voltage expressed in the stator reference frame.
$u_{si}(t)$	Stator voltage per phase.
$\bar{u}_s$	Space phasor of the stator voltage expressed in the stator reference frame.
$\bar{u}_s'$	Space phasor of the stator voltage expressed in the rotor reference frame.
ρr	Phase angle of the rotor flux linkage space phasor with respect to the direct-axis of the stator reference frame.
ρs	Phase angle of the stator flux linkage space phasor with respect to the direct-axis of the stator reference frame.
θ_m	Stator to rotor angle.
θ_r	Rotor angle.
θ_s	Stator angle.
$\Psi_{ri}(t)$	Flux linkage per rotor winding.
$\bar{\Psi}_r$	Space phasor of the rotor flux linkage expressed in the rotor reference frame.
$\bar{\Psi}_r'$	Space phasor of the rotor flux linkage expressed in the stator reference frame.
$\Psi_{si}(t)$	Flux linkage per stator winding.
$\bar{\Psi}_s$	Space phasor of the stator flux linkage expressed in the stator reference frame.
$\bar{\Psi}_s'$	Space phasor of the stator flux linkage expressed in the rotor reference frame.
λ'^{e}_{dr}	*d-axis* rotor flux

Subscripts.

α/β	Direct- and quadrature-axis components in the rotor reference frame.
d/q	Rotor direct- and quadrature-axis components in the stator reference frame.
D/Q	Stator direct and quadrature-axis components in the stator reference frame.
g	General reference frame.
m	Magnetizing.
r	Rotor.
ra, rb, rc	Rotor phases.
Ref	Reference.
s	Stator.
sA, sB, sC	Stator phases.
x/y	Direct- and quadrature-axis components in general reference frame or in special reference frames.

Superscripts

e	synchronous reference frame
r	rotor reference fame
s	stationary reference frame

CHAPTER 1

INTRODUCTION

1.1 Historical Review

The history of electrical motors goes back as far as 1820, when Hans Christian Oersted discovered the magnetic effect of an electric current. One year later, Michael Faraday discovered the electromagnetic rotation and built the first primitive D.C. motor. Faraday went on to discover electromagnetic induction in 1831, but it was not until 1883 that Tesla invented the A.C asynchronous motor.

Currently, the main types of electric motors are still the same, DC, AC asynchronous and synchronous, all based on Oersted, Faraday and Tesla's theories developed and discovered more than a hundred years ago.

Since its invention, the AC asynchronous motor, also named induction motor, has become the most widespread electrical motor in use today. At present, 67% of all the electrical energy generated in the UK is converted to mechanical energy for utilization. In Europe the electrical drives business is worth approximately $1.0 Billion/ Annum.

These facts are due to the induction motors advantages over the rest of motors. The main advantage is that induction motors do not require an electrical connection between stationary and rotating parts of the motor. Therefore, they do not need any mechanical commutator (brushes), leading to the fact that they are maintenance free motors. Induction motors also have low weight and inertia, high efficiency and a high overload capability. Therefore, they are cheaper and more robust, and less prove to any failure at high speeds. Furthermore, the motor can work in explosive environments because no sparks are produced.

Taking into account all the advantages outlined above, induction motors must be considered the perfect electrical to mechanical energy converter. However, mechanical energy is more than often required at variable speeds, where the speed control system is not a trivial matter.

The only effective way of producing an infinitely variable induction motor speed drive is to supply the induction motor with three phase voltages of variable frequency and variable amplitude. A variable frequency is required because the rotor speed depends on the speed of the rotating magnetic field provided by the stator. A variable voltage is required because the motor impedance reduces at low frequencies and consequently the current has to be limited by means of reducing the supply voltages [2] [1].

Before the days of power electronics, a limited speed control of induction motor was achieved by switching the three-stator windings from delta connection to star connection, allowing the voltage at the motor windings to be reduced. Induction motors are also available with more than three stator windings to allow a change of the number of pole pairs. However, a motor with several windings is more expensive because more than three connections to the motor are needed and only certain discrete speeds are available. Another alternative method of speed control can be realized by means of a wound rotor induction motor, where the rotor winding ends are brought out to slip rings. However, this method obviously removes most of the advantages of induction motors and it also introduces additional losses. By connecting resistors or reactance's in series with the stator windings of the induction motors, poor performance is achieved.

At that time the above described methods were the only ones available to control the speed of induction motors, whereas infinitely variable speed drives with good performances for DC motors already existed. These drives not only permitted the operation in four quadrants but also covered a wide power range. Moreover, they had a good efficiency, and with a suitable control even a good dynamic response. However, its main drawback was the compulsory requirement of brushes [1].

With the enormous advances made in semiconductor technology during the last 20 years, the required conditions for developing a proper induction motor drive are present. These conditions can be divided mainly in two groups:

- The decreasing cost and improved performance in power electronic switching devices.
- The possibility of implementing complex algorithms in the new microprocessors

However, one precondition had to be made, which was the development of suitable methods to control the speed of induction motors, because in contrast to its mechanical simplicity their complexity regarding their mathematical structure (multivariable and non-linear) is not a trivial matter.

It is in this field, that considerable research effort is devoted. The aim being to find even simpler methods of speed control for induction machines. One method, which is popular at the moment, is Vector Control [3].

1.2 Literature Review

Vector or Field-Oriented Control: As mentioned before, ideally a vector-controlled induction motor drive behaves like a separately excited de motor drive. The vector control is also known as decoupling, orthogonal, or trans vector control. It is a revolutionary invention in ac drives Technology. The higher order and coupling model of the machine that gives complex stability and sluggish response problems in a scalar controlled drive tend to vanish with vector control. Vector control techniques can be classified as indirect or feed forward method and direct or feedback method depending on the method of unit vector generation for vector rotation. There is also classification of control based on orientation with air gap flux, rotor flux, stator flux. [26], [32], [40]-[43]

Indirect or Feedforward Vector Control: K Hasse is the inventor of indirect vector control which he submitted for his doctoral thesis in 1969 in the Tech. Univ. The synchronously rotating Cartesian vector components of stator current i_{qs} and i_{ds} are controlled independently to control the torque and rotor flux, respectively. The airgap or stator flux orientation is also possible, but it can be shown that rotor flux orientation gives the true decoupling control. It is feed forward method in which cosine and sine vectors are generated in feed forward manner [15],[16],[44].

Direct or Feedback vector control: F. Blaschke invented the direct vector control method in 1971 , and later he submitted it for his doctoral thesis in the Tech. univ. Of Braunschweig in 1974. This is similar to the indirect or feed forward vector control but in this the unit vectors are generated by the feedback manner i.e. by sensing the rotor flux with the help of the sensors.[47].

Direct Torque Control: Takahashi and Depenbrock proposed a-high performance scalar control method which is popularly known as direct self control (DSC *)* or direct torque control (DTC). It can be shown that the developed torque of a machine is proportional to the product of synchronously rotating stator flux, rotor flux and the angle between them In a PWM inverter-fed machine.[24][25]

Adaptive and Optimal Control: A classical control design based on linear plant model and time invariant parameters can hardly be accepted for high performance control A vector-controlled drive can give fast response by decoupling control, but its poles and zeros can vary due to plant parameter (electrical and mechanical) variation. The electrical parameters of a machine may vary by saturation, temperature and skin effect, and the mechanical parameters are determined by the coupling 1oad.A high gain negative feedback loop can linearize and attenuate parameter variation and external disturbance, In adaptive control, the controller parameters (and sometimes the structure) vary to adapt continuously the variation of plant parameters to give the desired stability, dead-beat response and robustness[64]

Self-Tuning Control: The adaptive control an be classified as either explicit or implicit. A simple example of explicit or direct adaptive control is the gain scheduling control of an inertia (J) varying speed control system provided the parameter can be identified on real-time basis. In a more complex self-tuning control (STC), the system poles can be assigned uniquely, or poles, zeroes and gain may remain unique irrespective of parameter variations. In such a control system, a plant parameter estimation algorithm solves the plant model on-line by observer method. A tuning algorithm then adjusts the control parameters based on the estimation of plant parameters. The tuning of slip gain (Ks) of indirect vector control is another example of this method. Of course, the global stability of such a system should be assured.

Model Referencing : Adaptive Control the examples of implicit or indirect adaptive control methods are model referencing adaptive control (MRAC) and sliding mode control (SMC). Such a control gives robust performance of the drive, i.e., the response is not affected by any parameter variation (such as J), or load disturbance effect. In & (RAC), the plant response is forced to track the response of a reference model irrespective of plant parameter variation or load disturbance. Note that the reference model should be designed on the worst case basis. Otherwise, within the limited power handling capability of inverter-machine unit, the control will fail to follow the reference. Therefore, the desired robustness of the control system is obtained at the cost of optimum response speed. The MRAC is not generally preferred in the main drive control due to response chattering problem, But, it has found acceptance in feedback signal estimation and slip gain tuning of vector-controlled drive.

Sliding Mode Control: In sliding mode control, the "reference model" is stored in the form of predefined phase plane trajectory, and the drive system response is forced to follow or "slide" along the trajectory by a switching control algorithm. The structure or topology of the control is varied intentionally between the positive and negative feedback control modes so that the average response of the system is stable although in individual structure it may be unstable.

The disadvantages of sliding mode control are the chattering or oscillatory output, acoustic noise and , of course, the sub-optimal transient response. The chattering output may not be acceptable for precision control applications. The harmonic loss is enhanced to some extent. due to chattering. The chattering can be improved Kith higher PWM frequency, small sampling time of microprocessor and reduced time delay for computation of feedback signals Various chattering alleviation algorithms, such as low-pass filter in the forward path, elimination of the dither signal and transitioning to PI (proportionalintegral) control near steady state, hybriding with continuous state feedback control, transitioning from discontinuous to continuous control near the sliding plane, and directly controlling the power switches have been proposed. Performance enhancement by adaptation off control parameters, trajectory boundary optimization and fuzzy sliding mode control have also been proposed.

Intelligent Control: Intelligent control is based on artificial intelligence (AI) which can be defined as computer emulation of human thinking process. The AI techniques are generally classified as expert system, Fuzzy logic, artificial neural network and genetic algorithm. Expert system, based on Boolean algebra, uses hard or precise computation, whereas fuzzy logic, neural network and genetic algorithm use soft or approximate computation. With a control based on *AI,* a system is said to be “intelligent”, “autonomous”, “adaptive”, self organizing” or “learning control”. The conventional control design is based on mathematical model of the plant. Often the plant model is unknown, or ill-defined. Or*,* the system may be nonlinear, complex, multivariable with parameter variation problem. An intelligent control system can identify the model, if necessary, and give predicted performance even with wide range of parameter variation. Of course, if R model is available; it can be used for simulation study where the control can be optimized by iteration.

Fuzzy Logic Control: Fuzzy logic, unlike Boolean logic, deals with problems that have vagueness, uncertainty, or imprecision, and uses membership functions with values between 0 and 1 to solve the problem Fuzzy control, similar to expert system based control, is described by a set of IF ... THEN production rules, and is often defined as fuzzy expert system.[27]-[31]

1.3 Structure of the thesis.

The work presented in this thesis is organized in four main chapters. Theses five chapters are Structured as follows.

Chapter 1 In this chapter the introduction to the drives and the literature review is of the vector control is presented.

Chapter 2 is entitled “Induction Motor Modeling.”. It introduces a mathematical model of cage rotor induction motors. Different ways of implementing these models are presented as well as some simulations corroborating its validity. It must be said that all simulations are obtained from MATLAB/Simulink. The elements of space phasor notation are also introduced and used to develop a compact notation.

Chapter 3 is entitled “Vector Control”. In this chapter discussed topics are types of induction motor controls briefly and the vector control in depth that too modeling of indirect vector control. i.e. designing of indirect vector control block in the SIMULINK of the Matlab

Chapter 4 is entitled “Fuzzy Logic Controller”. This chapter is devoted to introduce the Artificial Intelligence technique i.e. fuzzy logic. What is Fuzzy logic and how is it is different from other conventional methods, applicability of fuzzy logic in engineering applications and then the application of fuzzy logic controller how it is designed , rules used.

Chapter 5 is titled as “Simulink Model” in this chapter the Simulink modeling is described and the algorithm and flowchart for the calculation of parameters is described

Chapter 6 is entitled as “Results and discussion” this chapter is focused on the discussion on the obtained results, i.e. transient torque improvement, rotor flux stabilization.

Chapter 7, entitled "Conclusions and Future work", all achievements are summarized and appropriate conclusions are drawn and future work is also included.

1.4 Aims of the thesis

The present thesis deals with the development of a High Precision Servo Control of Induction Motor Drive using Indirect Vector Control with Fuzzy PI Controller. This fuzzy Controller that has improved performance compared to the proposed model is [65]. The main improvement has been the transient torque reduction both in magnitude and time during starting. Also the rapid stabilization of rotor flux. The overall performance of the drive system to be excellent.

CHAPTER 2

INDUCTION MOTOR MODELING

2.1 Introduction.

A dynamic model of the machine subjected to control must be known in order to understand and design vector controlled drives. Due to the fact that every good control has to face any possible change of the plant, it could be said that the dynamic model of the machine could be just a good approximation of the real plant. Nevertheless, the model should incorporate all the important dynamic effects occurring during both steady-state and transient operations. Furthermore, it should be valid for any changes in the inverter's supply such as voltages or currents.

Such a model can be obtained by means of either the space vector phasor theory or two-axis theory of electrical machines. Despite the compactness and the simplicity of the space phasor theory, both methods are actually close and both methods will be explained.

For simplicity, the induction motor considered will have the following assumptions:

- Symmetrical two-pole, three phases windings.
- The slotting effects are neglected.
- The permeability of the iron parts is infinite.
- The flux density is radial in the air gap.
- Iron losses are neglected.
- The stator and the rotor windings are simplified as a single, multi-turn full pitch coil situated on the two sides of the air gap.

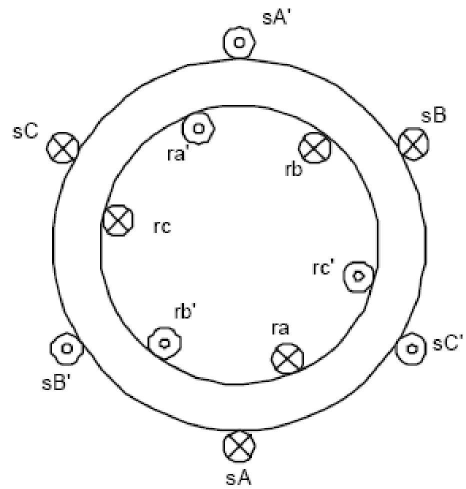


Figure 2.1. Cross-section of an elementary symmetrical three-phase machine.

2.2 Voltage equations.

The stator voltages will be formulated in this section from the motor natural frame, which is the stationary reference frame fixed to the stator. In a similar way, the rotor voltages will be formulated to the rotating frame fixed to the rotor. In the stationary reference frame, the equations can be expressed as follows:

$$u_{sA}(t) = R_s i_{sA}(t) + \frac{d\Psi_{sA}(t)}{dt} \tag{2.1}$$

$$u_{sB}(t) = R_s i_{sB}(t) + \frac{d\Psi_{sB}(t)}{dt} \tag{2.2}$$

$$u_{sC}(t) = R_s i_{sC}(t) + \frac{d\Psi_{sC}(t)}{dt} \tag{2.3}$$

Similar expressions can be obtained for the rotor:

$$u_{ra}(t) = R_r i_{ra}(t) + \frac{d\Psi_{ra}(t)}{dt} \tag{2.4}$$

$$u_{rb}(t) = R_r i_{rb}(t) + \frac{d\Psi_{rb}(t)}{dt} \tag{2.5}$$

$$u_{rc}(t) = R_r i_{rc}(t) + \frac{d\Psi_{rc}(t)}{dt} \tag{2.6}$$

The instantaneous stator flux linkage values per phase can be expressed as:

$$\Psi_{sA} = \overline{L}_s i_{sA} + \overline{M}_s i_{sB} + \overline{M}_s i_{sC} + \overline{M}_{sr} \cos\theta_m i_{ra} + \overline{M}_{sr} \cos(\theta_m + 2\pi/3) i_{rb} + \overline{M}_{sr} \cos(\theta_m + 4\pi/3) i_{rc} \tag{2.7}$$

$$\Psi_{sB} = \overline{M}_s i_{sA} + \overline{L}_s i_{sB} + \overline{M}_s i_{sC} + \overline{M}_{sr} \cos(\theta_m + 4\pi/3) i_{ra} + \overline{M}_{sr} \cos\theta_m i_{rb} + \overline{M}_{sr} \cos(\theta_m + 2\pi/3) i_{rc} \tag{2.8}$$

$$\Psi_{sC} = \overline{M}_s i_{sA} + \overline{M}_s i_{sB} + \overline{L}_s i_{sC} + \overline{M}_{sr} \cos(\theta_m + 2\pi/3) i_{ra} + \overline{M}_{sr} \cos(\theta_m + 4\pi/3) i_{rb} + \overline{M}_{sr} \cos\theta_m i_{rc} \tag{2.9}$$

In a similar way, the rotor flux linkages can be expressed as follows:

$$\Psi_{ra} = \overline{M}_{sr} \cos(-\theta_m) i_{sA} + \overline{M}_{sr} \cos(-\theta_m + 2\pi/3) i_{sB} + \overline{M}_{sr} \cos(-\theta_m + 4\pi/3) i_{sC} + \overline{L}_r i_{ra} + \overline{M}_r i_{rb} + \overline{M}_r i_{rc} \tag{2.10}$$

$$\Psi_{rb} = \overline{M}_{sr} \cos(-\theta_m + 4\pi/3) i_{sA} + \overline{M}_{sr} \cos(-\theta_m) i_{sB} + \overline{M}_{sr} \cos(-\theta_m + 2\pi/3) i_{sC} + \overline{M}_r i_{ra} + \overline{L}_r i_{rb} + \overline{M}_r i_{rc} \tag{2.11}$$

$$\Psi_{rc} = \overline{M}_{sr} \cos(-\theta_m + 2\pi/3) i_{sA} + \overline{M}_{sr} \cos(-\theta_m + 4\pi/3) i_{sB} + \overline{M}_{sr} \cos(-\theta_m) i_{sC} + \overline{M}_r i_{ra} + \overline{L}_r i_{rb} + \overline{M}_r i_{rc} \tag{2.12}$$

Taking into account all the previous equations, and using the matrix notation in order to compact all the expressions, the following expression is obtained:

$$\begin{bmatrix} u_{sA} \\ u_{sB} \\ u_{sC} \\ u_{ra} \\ u_{rb} \\ u_{rc} \end{bmatrix} = \begin{bmatrix} R_s + p\overline{L}_s & p\overline{M}_s & p\overline{M}_s & p\overline{M}_{sr}\cos\theta_m & p\overline{M}_{sr}\cos\theta_{m1} & p\overline{M}_{sr}\cos\theta_{m2} \\ p\overline{M}_s & R_s + p\overline{L}_s & p\overline{M}_s & p\overline{M}_{sr}\cos\theta_{m2} & p\overline{M}_{sr}\cos\theta_m & p\overline{M}_{sr}\cos\theta_{m1} \\ p\overline{M}_s & p\overline{M}_s & R_s + p\overline{L}_s & p\overline{M}_{sr}\cos\theta_{m1} & p\overline{M}_{sr}\cos\theta_{m2} & p\overline{M}_{sr}\cos\theta_m \\ p\overline{M}_{sr}\cos\theta_m & p\overline{M}_{sr}\cos\theta_{m1} & p\overline{M}_{sr}\cos\theta_{m2} & R_r + p\overline{L}_r & p\overline{M}_r & p\overline{M}_r \\ p\overline{M}_{sr}\cos\theta_{m2} & p\overline{M}_{sr}\cos\theta_m & p\overline{M}_{sr}\cos\theta_{m1} & p\overline{M}_r & R_r + p\overline{L}_r & p\overline{M}_r \\ p\overline{M}_{sr}\cos\theta_{m1} & p\overline{M}_{sr}\cos\theta_{m2} & p\overline{M}_{sr}\cos\theta_m & p\overline{M}_r & p\overline{M}_r & R_r + p\overline{L}_r \end{bmatrix} \cdot \begin{bmatrix} i_{sA} \\ i_{sB} \\ i_{sC} \\ i_{ra} \\ i_{rb} \\ i_{rc} \end{bmatrix} \tag{2.13}$$

2.2.1 Park's transform.

In order to reduce the expressions of the induction motor equation voltages given in equation 2.1 to equation 2.6 and obtain constant coefficients in the differential equations, the Park's transform will be applied. Physically, it can be understood as transforming the three windings of the induction motor to just two windings, as it is

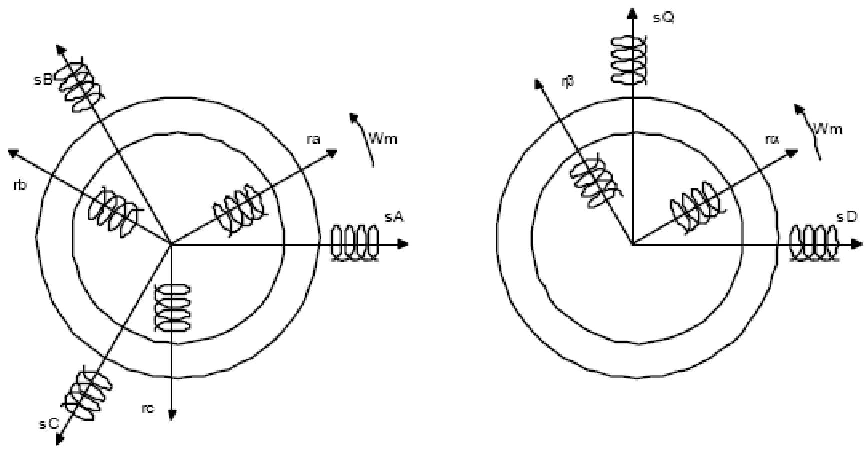


Figure 2.2 Schema of the equivalence physics transformation.

In the symmetrical three-phase machine, the direct- and the quadrature-axis stator magnitudes are fictitious. The equivalencies for these direct (D) and quadrature (Q) magnitudes with the magnitudes per phase are as follows:

$$\begin{bmatrix} u_{s0} \\ u_{sD} \\ u_{sQ} \end{bmatrix} = c. \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ \cos\theta & \cos(\theta - 2\pi/3) & \cos(\theta + 2\pi/3) \\ -\sin\theta & -\sin(\theta - 2\pi/3) & -\sin(\theta + 2\pi/3) \end{bmatrix} . \begin{bmatrix} u_{sA} \\ u_{sB} \\ u_{sC} \end{bmatrix} \tag{2.14}$$

$$\begin{bmatrix} u_{sA} \\ u_{sB} \\ u_{sC} \end{bmatrix} = c_1. \begin{bmatrix} 1/\sqrt{2} & \cos\theta & -\sin\theta \\ 1/\sqrt{2} & \cos(\theta - 2\pi/3) & -\sin(\theta - 2\pi/3) \\ 1/\sqrt{2} & \cos(\theta + 2\pi/3) & -\sin(\theta + 2\pi/3) \end{bmatrix} . \begin{bmatrix} u_{s0} \\ u_{sD} \\ u_{sQ} \end{bmatrix} \tag{2.15}$$

Where c, c_1 are the constants with the values 2/3 or 1 for the so-called non-power invariant form or the value $\sqrt{2/3}$ for the power-invariant form as it is explained in section 2.3.3. These previous equations can be applied as well for any other magnitudes such as currents and fluxes.

Notice how the expression 2.13 can be simplified into a much smaller expression in 2.16 by means of applying the mentioned Park's transform.

$$\begin{bmatrix} u_{sD} \\ u_{sQ} \\ u_{r\alpha} \\ u_{r\beta} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -L_s p\theta_s & pL_m & -L_m(P.w_m + p\theta_r) \\ L_s p\theta_s & R_s + pL_s & L_m(P.w_m + p\theta_r) & pL_m \\ pL_m & -L_m(p\theta_s - P.w_m) & R_r + pL_r & -L_r p\theta_r \\ L_m(p\theta_s - P.w_m) & pL_m & L_r p\theta_r & R_r + pL_r \end{bmatrix} . \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} \tag{2.16}$$

Where $L_s = \overline{L}_s - \overline{M}_s$, $L_r = \overline{L}_r - \overline{M}_r$ and $L_m = \frac{3}{2}\overline{M}_{sr}$.

2.2.2 Voltage matrix equations.

If the matrix expression 2.16 is simplified, new matrixes are obtained as shown in equations 2.17, 2.18 and 2.19.

2.2.2.1 – Fixed to the stator.

It means that $w_s = 0$ and consequently $w_r = -w_m$.

$$\begin{bmatrix} u_{sD} \\ u_{sQ} \\ u_{rd} \\ u_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & 0 & pL_m & 0 \\ 0 & R_s + pL_s & 0 & pL_m \\ pL_m & -P.w_m L_m & R_r + pL_r & P.w_m L_r \\ -P.w_m L_m & pL_m & -P.w_m L_r & R_r + pL_r \end{bmatrix} . \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{rd} \\ i_{rq} \end{bmatrix} \tag{2.17}$$

2.2.2.2 – Fixed to the rotor.

It means that $w_r = 0$ and consequently $w_s = w_m$.

$$\begin{bmatrix} u_{sD} \\ u_{sQ} \\ u_{rd} \\ u_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & 0 & pL_m & 0 \\ 0 & R_s + pL_s & 0 & pL_m \\ pL_m & -P.w_m L_m & R_r + pL_r & P.w_m L_r \\ -P.w_m L_m & pL_m & -P.w_m L_r & R_r + pL_r \end{bmatrix} . \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{rd} \\ i_{rq} \end{bmatrix} \tag{2.18}$$

2.2.2.3 – Fixed to the synchronism.

It means that $w_r = sw_s$.

$$\begin{bmatrix} u_{sD} \\ u_{sQ} \\ u_{rd} \\ u_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -L_s w_s & pL_m & -L_s w_s \\ L_s w_s & R_s + pL_s & L_m w_s & pL_m \\ pL_m & -L_m s w_s & R_r + pL_r & -L_r s w_s \\ L_m s w_s & pL_m & L_r s w_s & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad (2.19)$$

2.3 Space phasor notation.

Space phasor notation allows the transformation of the natural instantaneous values of a three phase system onto a complex plane located in the cross section of the motor. In this plane, the space phasor rotate with an angular speed equal to thewherein angular frequency of the three phase supply system. A space phasor rotating with the same angular speed, for example, can describe the rotating magnetic field.

Moreover, in the special case of the steady state, where the supply voltage is sinusoidal and symmetric, the space phasor become equal to three-phase voltage phasors, allowing the analysis in terms of complex algebra. It is shown in figure 2.3 the equivalent schematic for this new model

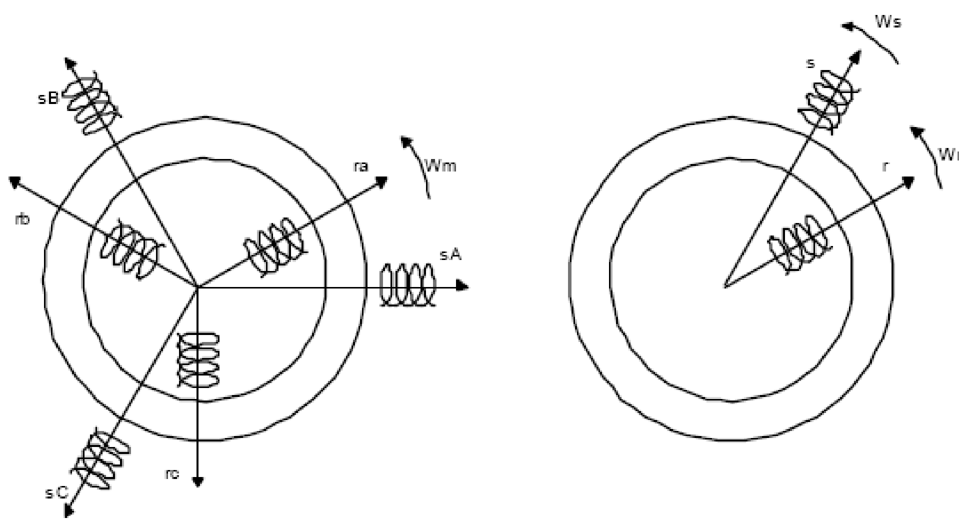


Figure 2.3. On the right the equivalent two rotating windings induction motor.

In order to transform the induction motor model, in natural co-ordinates, into its equivalent space phasor form, the 120º operator is introduced:

$$a = e^{j2\pi/3}$$
$$a^2 = e^{j4\pi/3} \tag{2.20}$$

Thus, the current stator space phasor can be expressed as follows:

$$\bar{i}_s = c.\left[1 i_{sA}(t) + a i_{sB}(t) + a^2 i_{sC}(t)\right] \tag{2.21}$$

The factor "c", takes usually one of two different values either $2/3$ or $\sqrt{2/3}$. The factor $2/3$ makes the amplitude of any space phasor, which represents a three phase balanced system, equal to the amplitudes of one phase of the three-phase system. The factor $\sqrt{2/3}$ may also be used to define the power invariance of a three-phase system with its equivalent two-phase system (see section 2.3.3).

2.3.1 Current space phasors.

During this section the induction machine assumptions introduced in the section 2.1 will be further considered. It is represented in figure 2.4 the model of the induction machine with two different frames, the D-Q axis which represent the stationary frame fixed to the stator, and the α-β axis which represent rotating frame fixed to the rotor.

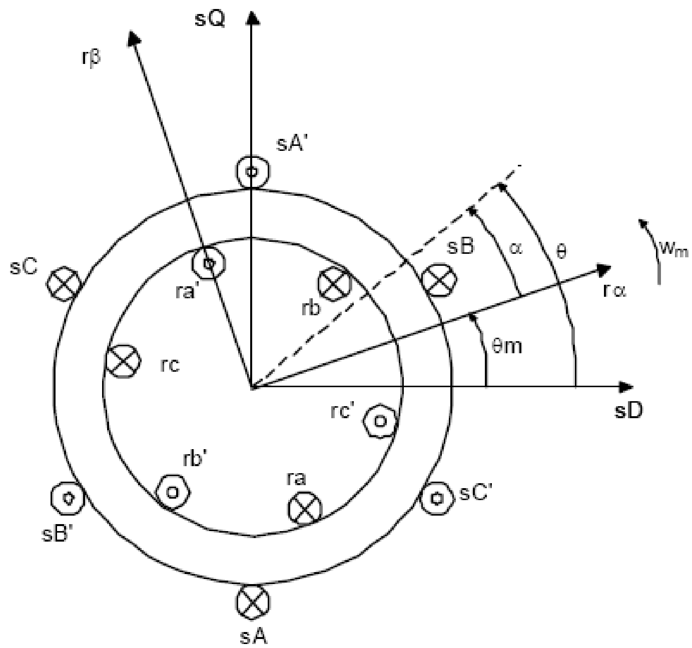


Figure 2.4. Cross-section of an elementary symmetrical three-phase machine, with two different frames, the D-Q axis which represent the stationary frame fixed to the stator, and α-β axis which represent rotating frame fixed to the rotor.

The stator current space phasor can be expressed as follows:

$$\bar{i}_s = \frac{2}{3}\left[i_{sA}(t) + a i_{sB}(t) + a^2 i_{sC}(t) \right] = \left| \bar{i}_s \right| e^{i\theta} \qquad (2.22)$$

Expressed in the reference frame fixed to the stator, the real-axis of this reference frame is denoted by sD and its imaginary-axis by sQ.

The equivalence between the stator phasor and the D-Q two-axis components is as follows:

$$\bar{i}_s = i_{sD}(t) + j i_{sQ}(t) \qquad (2.23)$$

or:

$$\mathrm{Re}(\bar{i}_s) = \mathrm{Re}\left[\frac{2}{3}(i_{sA} + ai_{sB} + a^2 i_{sC})\right] = i_{sD}$$

$$\mathrm{Im}(\bar{i}_s) = \mathrm{Im}\left[\frac{2}{3}(i_{sA} + ai_{sB} + a^2 i_{sC})\right] = i_{sQ} \tag{2.24}$$

The relationship between the space phasor current and the real stator phase currents can be expressed as follows:

$$\mathrm{Re}(\bar{i}_s) = \mathrm{Re}\left[\frac{2}{3}(i_{sA} + ai_{sB} + a^2 i_{sC})\right] = i_{sA}$$

$$\mathrm{Re}(a^2\bar{i}_s) = \mathrm{Re}\left[\frac{2}{3}(a^2 i_{sA} + i_{sB} + ai_{sC})\right] = i_{sB}$$

$$\mathrm{Re}(a\bar{i}_s) = \mathrm{Re}\left[\frac{2}{3}(ai_{sA} + a^2 i_{sB} + i_{sC})\right] = i_{sC} \tag{2.25}$$

In a similar way, the space phasor of the rotor current can be written as follows:

$$\bar{i}_r = \frac{2}{3}\left[i_{ra}(t) + a.i_{rb}(t) + a^2.i_{rc}(t)\right] = \left|\bar{i}_r\right| e^{j\alpha} \tag{2.26}$$

Expressed in the reference frame fixed to the rotor, the real-axis of this reference frame is denoted by rα and its imaginary-axis by rβ.

The space phasor of the rotor current expressed in the stationary reference frame fixed to the stator can be expressed as follows:

$$\overline{i_r} = \left|\overline{i_r}\right| e^{j\theta} = \left|\overline{i_r}\right| e^{j(\alpha+\theta_m)} \tag{2.27}$$

The equivalence between the current rotor space phasor and the α-β two-axis is as follows:

$$\overline{i_r} = i_{r\alpha}(t) + ji_{r\beta}(t) \tag{2.28}$$

Or

$$\mathrm{Re}(\bar{i}_r) = \mathrm{Re}\left[\frac{2}{3}(i_{ra} + a i_{rb} + a^2 i_{rc})\right] = i_{r\alpha}$$

$$\mathrm{Im}(\bar{i}_r) = \mathrm{Im}\left[\frac{2}{3}(i_{ra} + a i_{rb} + a^2 i_{rc})\right] = i_{r\beta} \tag{2.29}$$

The relationship between the space phasor current and the real stator currents can be expressed as follows:

$$\mathrm{Re}(\bar{i}_r) = \mathrm{Re}\left[\frac{2}{3}(i_{ra} + a i_{rb} + a^2 i_{rc})\right] = i_{ra}$$

$$\mathrm{Re}(a^2\bar{i}_r) = \mathrm{Re}\left[\frac{2}{3}(a^2 i_{ra} + i_{rb} + a i_{rc})\right] = i_{rb}$$

$$\mathrm{Re}(a\bar{i}_r) = \mathrm{Re}\left[\frac{2}{3}(a i_{ra} + a^2 i_{rb} + i_{rc})\right] = i_{rc} \tag{2.30}$$

The magnetizing current space-phasor expressed in the stationary reference frame fixed to the stator can be obtained as follows:

$$\overline{i_m} = \overline{i_s} + \left({}^{N_{re}}\!/\!_{N_{se}}\right)\overline{i'_r} \tag{2.31}$$

2.3.2 Flux linkage space phasor.

In this section the flux linkages will be formulated in the stator phasor notation according to different reference frames.

2.3.2.1- Stator flux-linkage space phasor in the stationary reference frame fixed to the stator. Similarly to the definitions of the stator current and rotor current space phasors, it is possible to define a space phasor for the flux linkage as follows:

$$\overline{\Psi}_s = \frac{2}{3}(\Psi_{sA} + a\Psi_{sB} + a^2\Psi_{sC}) \tag{2.32}$$

If the flux linkage equations 2.7, 2.8, 2.9 are substituted in equation 2.32, the space phasor for the stator flux linkage can be expressed as follows:

$$\overline{\Psi}_s = \frac{2}{3}\begin{bmatrix} i_{sA}(\overline{L}_s + a\overline{M}_s + a^2\overline{M}_s) + i_{sB}(\overline{M}_s + a\overline{L}_s + a^2\overline{M}_s) + i_{sC}(\overline{M}_s + a\overline{M}_s + a^2\overline{L}_s) + \\ i_{ra}(\overline{M}_{sr}\cos\theta_m + a\overline{M}_{sr}\cos(\theta_m + 4\pi/3) + a^2\overline{M}_{sr}\cos(\theta_m + 2\pi/3)) + \\ i_{rb}(\overline{M}_{sr}\cos(\theta_m + 2\pi/3) + a\overline{M}_{sr}\cos\theta_m + a^2\overline{M}_{sr}\cos(\theta_m + 4\pi/3)) + \\ i_{rc}(\overline{M}_{sr}\cos(\theta_m + 4\pi/3) + a\overline{M}_{sr}\cos(\theta_m + 2\pi/3) + a^2\overline{M}_{sr}\cos\theta_m) \end{bmatrix} \tag{2.33}$$

Developing the previous expression 2.33, it is obtained the following expression:

$$\overline{\Psi}_s = \frac{2}{3}\begin{bmatrix} i_{sA}(\overline{L}_s + a\overline{M}_s + a^2\overline{M}_s) + ai_{sB}(a^2\overline{M}_s + \overline{L}_s + a\overline{M}_s) + a^2 i_{sC}(a\overline{M}_s + a^2\overline{M}_s + \overline{L}_s) + \\ i_{ra}(\overline{M}_{sr}\cos\theta_m + a\overline{M}_{sr}\cos(\theta_m + 4\pi/3) + a^2\overline{M}_{sr}\cos(\theta_m + 2\pi/3)) + \\ i_{rb}(\overline{M}_{sr}\cos(\theta_m + 2\pi/3) + a\overline{M}_{sr}\cos\theta_m + a^2\overline{M}_{sr}\cos(\theta_m + 4\pi/3)) + \\ i_{rc}(\overline{M}_{sr}\cos(\theta_m + 4\pi/3) + a\overline{M}_{sr}\cos(\theta_m + 2\pi/3) + a^2\overline{M}_{sr}\cos\theta_m) \end{bmatrix} \tag{2.34}$$

And finally, expression 2.34 can be represented as follows:

$$\begin{aligned}\Psi_s &= \bar{i}_s(\overline{L}_s + a\overline{M}_s + a^2\overline{M}_s) + \bar{i}_r(\overline{M}_{sr}\cos\theta_m + a\overline{M}_{sr}\cos(\theta_m + 4\pi/3) + a^2\overline{M}_{sr}\cos(\theta_m + 2\pi/3)) \\ &= (\overline{L_s} - \overline{M_s})\bar{i}_s + 1.5\cos\theta_m\overline{M}_{sr}\bar{i}_r \\ &= (\overline{L_s} - \overline{M_s})\overline{i_s} + 1.5\overline{M}_{sr}\overline{i_r}e^{j\theta_m} \\ &= (\overline{L_s} - \overline{M_s})\overline{i_s} + 1.5\overline{M}_{sr}\bar{i}_r{}' \\ &= L_s\overline{i_s} + L_m\bar{i}_r{}' \end{aligned} \tag{2.35}$$

Where L_s is the total three-phase stator inductance and L_m is the so-called three-phase magnetizing inductance. Finally, the space phasor of the flux linkage in the stator depends on two components, being the stator currents and the rotor currents. Once more, the flux linkage magnitude can be expressed in two-axis as follows:

$$\overline{\Psi}_s = \Psi_{sD} + j\Psi_{sQ} \tag{2.36}$$

Where L_r is the total three-phase rotor inductance and L_m is the so-called three-phase magnetizing inductance. $\bar{i}_s'$ is the stator current space phasor expressed in the frame fixed to the rotor.

Once more the flux linkage magnitude can be expressed in the two-axis form as follows:

$$\overline{\Psi}_r = \Psi_{r\alpha} + j\Psi_{r\beta} \tag{2.37}$$

Where its direct component is equal to:

$$\Psi_{r\alpha} = L_r i_{r\alpha} + L_m i_{s\alpha} \tag{2.38}$$

And its quadrature component is expressed as:

$$\Psi_{r\beta} = L_r i_{r\beta} + L_m i_{s\beta} \tag{2.39}$$

2.3.2.2- Rotor flux-linkage space phasor in the stationary reference frame fixed to the stator. The rotor flux linkage can also be expressed in the stationary reference frame using the previously introduced transformation $e^{j\theta_m}$, and can be written as:

$$\begin{aligned}\overline{\Psi'}_r &= \Psi_{rd} + j\Psi_{rq} \\ &= \overline{\Psi_r} e^{j\theta_m} \\ &= (\Psi_{r\alpha} + j\Psi_{r\beta}) e^{j\theta_m}\end{aligned} \tag{2.40}$$

The space phasor of the rotor flux linkage can be expressed according to the fixed coordinates as follows:

$$\overline{\Psi}'_r = L_r \bar{i}'_r + L_m \bar{i}'_s e^{j\theta_m}$$

$$= L_r \bar{i}'_r + L_m \bar{i}_s \tag{2.41}$$

The relationship between the stator current referred to the stationary frame fixed to the stator and the rotational frame fixed to the rotor is as follows:

$$\begin{aligned} \bar{i}_s &= \bar{i}'_s e^{j\theta_m} \\ \bar{i}'_s &= \bar{i}_s e^{-j\theta_m} \end{aligned} \tag{2.42}$$

Where

$$\begin{aligned} \bar{i}_s &= i_{sd} + j i_{sq} \\ \bar{i}'_s &= i_{s\alpha} + j i_{s\beta} \end{aligned} \tag{2.43}$$

From figure 2.5, the following equivalencies can be deduced:

$$\begin{aligned} \bar{i}_s &= \left|\bar{i}_s\right| e^{j\theta} \\ \bar{i}'_s &= \left|\bar{i}'_s\right| e^{j\alpha} \\ \bar{i}'_s &= \left|\bar{i}'_s\right| e^{j(\theta-\theta_m)} \\ \bar{i}'_s &= \bar{i}_s e^{-j\theta_m} \end{aligned} \tag{2.44}$$

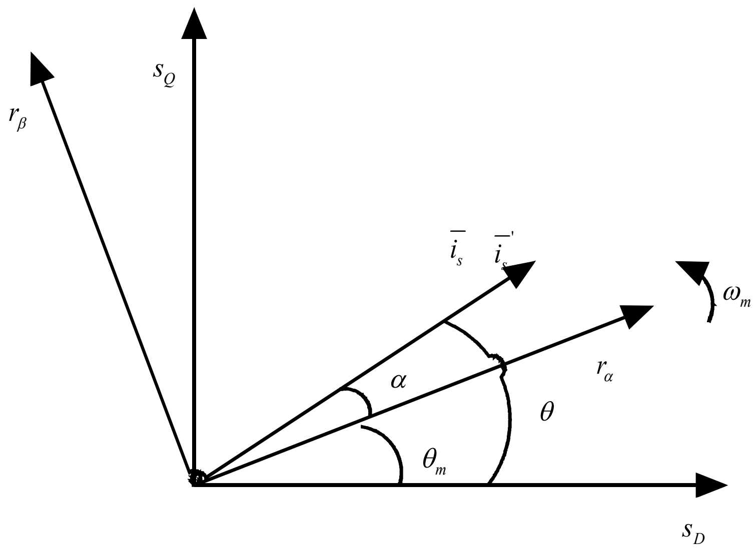


Figure 2.5. Stator-current space phasor expressed in accordance with the rotational frame fixed to the rotor and the stationary frame fixed to the stator.

$$\overline{\Psi}'_s = \overline{\Psi}_s e^{-j\theta_m}$$

$$\overline{\Psi}'_s = \left(L_s \bar{i}_s + L_m \bar{i}'_r \right) e^{-j\theta_m}$$

$$\overline{\Psi}'_s = L_s \bar{i}_s + L_m \bar{i}_r \tag{2.45}$$

2.3.3. The space phasors of stator and rotor voltages.

The space phasors for the stator and rotor voltages can be defined in a similar way like the one used for other magnitudes.

$$\begin{aligned} \bar{u}_s &= \frac{2}{3}\left[u_{sA}(t) + a.u_{sB}(t) + a^2.u_{sC}(t) \right] \\ &= u_{sD} + ju_{sQ} \\ &= \frac{2}{3}\left(u_{sA} - \frac{1}{2}u_{sB} - \frac{1}{2}u_{sC} \right) + j\frac{1}{\sqrt{3}}\left(u_{sB} - u_{sC} \right) \end{aligned} \tag{2.46}$$

$$\bar{u}_r = \frac{2}{3}\left[u_{ra}(t) + a.u_{rb}(t) + a^2.u_{rc}(t)\right]$$
$$= u_{r\alpha} + ju_{r\beta}$$
$$= \frac{2}{3}\left(u_{ra} - \frac{1}{2}u_{rb} - \frac{1}{2}u_{rc}\right) + j\frac{1}{\sqrt{3}}\left(u_{rb} - u_{rc}\right) \tag{2.47}$$

Where the stator voltage space phasor is referred to the stator stationary frame and the rotor voltage space phasor is referred to the rotating frame fixed to the rotor. Provided the zero component is zero ,it can also be said that:

$$u_{sA} = \mathrm{Re}\left(\bar{u}_s\right)$$
$$u_{sB} = \mathrm{Re}\left(a^2\bar{u}_s\right)$$
$$u_{sC} = \mathrm{Re}\left(a\bar{u}_s\right) \tag{2.48}$$

Equivalent expressions can also be obtained for the rotor.

2.3.4 Space-phasor form of the motor equations.

The space phasor forms of the voltage equations of the three-phase and quadrature-phase smooth air-gap machines will be presented. Firstly, the equations will be expressed in a general rotating reference frame, which rotates at a general speed w_g, and then to the references frames fixed to the stator, rotor and synchronous speed.

2.3.4.1 - Space-phasor voltage equations in the general reference frame.

If the vector in the figure 2.6 is the stator current, then its formulation in the space phasor form is as follows:

$$\bar{i}_{sg} = \bar{i}_s e^{-j\theta_g}$$
$$= i_{sx} + ji_{sy} \tag{2.49}$$

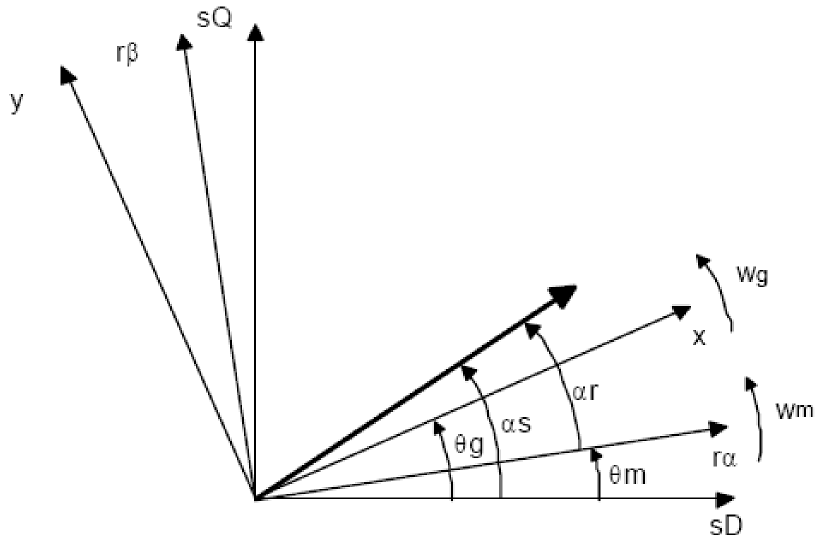


Figure 2.6. It is shown a magnitude represented by means of the vector, and its angle referred to the three different axis. The three different axis are: sD-sQ fixed to the stator, rα-rβ fixed to the rotor whose speed is wm, and finally the general frame represented by means of the axis x-y whose speed is equal to w_g.

In a similar way and for other magnitudes, it can be written the following equations:

$$\begin{aligned}\bar{u}_{sg} &= \bar{u}_s e^{-j\theta_g} \\ &= u_{sx} + ju_{sy}\end{aligned} \tag{2.50}$$

$$\begin{aligned}\overline{\Psi}_{sg} &= \overline{\Psi}_s e^{-j\theta_g} \\ &= \Psi_{sx} + j\Psi_{sy}\end{aligned} \tag{2.51}$$

Where the magnitudes are the voltage space phasor and the stator flux linkage respectively. However, if the magnitude in the figure 2.6 is for instance the rotor current, its space phasor notation will be:

$$\begin{aligned}\bar{i}_{rg} &= \bar{i}_r e^{-j(\theta_g - \theta_m)} \\ &= i_{sx} + ji_{sy}\end{aligned} \tag{2.52}$$

and for other magnitudes:

$$\bar{u}_{rg} = \bar{u}_r e^{-j(\theta_g - \theta_m)}$$
$$= u_{sx} + ju_{sy} \quad (2.53)$$

$$\overline{\Psi}_{rg} = \overline{\Psi}_r e^{-j(\theta_g - \theta_m)}$$
$$= \Psi_{sx} + j\Psi_{sy} \quad (2.54)$$

Manipulating the previous equations yields the following stator and rotor space phasor voltage equations in the general reference frame.

$$\bar{u}_{sg} e^{j\theta_g} = R_s \bar{i}_{sg} e^{j\theta_g} + \frac{d\left(\overline{\Psi}_{sg} e^{j\theta_g}\right)}{dt}$$
$$= R_s \bar{i}_{sg} e^{j\theta_g} + e^{j\theta_g} \frac{d\overline{\Psi}_{sg}}{dt} + je^{j\theta_g} w_g \overline{\Psi_{sg}} \quad (2.55)$$

$$\bar{u}_{rg} e^{j(\theta_g - \theta_m)} = R_r \bar{i}_{rg} e^{j(\theta_g - \theta_m)} + \frac{d\left(\overline{\Psi}_r e^{j(\theta_g - \theta_m)}\right)}{dt}$$
$$= R_r \bar{i}_{rg} e^{j(\theta_g - \theta_m)} + e^{j(\theta_g - \theta_m)} \frac{d\overline{\Psi}_{rg}}{dt} + je^{j(\theta_g - \theta_m)} \left(w_g - P.w_m\right) \overline{\Psi}_{rg} \quad (2.56)$$

Simplifying equation 2.55, it is obtained equation 2.56.

$$\bar{u}_{sg} = R_s \bar{i}_{sg} + \frac{d\overline{\Psi}_{sg}}{dt} + jw_g \overline{\Psi}_{sg} \quad (2.57)$$

$$\bar{u}_{rg} = R_r \bar{i}_{rg} + \frac{d\overline{\Psi}_{rg}}{dt} + j(w_g - Pw_m) \overline{\Psi}_{rg} \quad (2.58)$$

Where, the flux linkage space phasors are:

$$\overline{\Psi}_{sg} = L_s \bar{i}_{sg} + L_m \bar{i}_{rg}$$

$$\overline{\Psi}_{rg} = L_r \bar{i}_{rg} + L_m \bar{i}_{sg} \tag{2.59}$$

Using the two-axis notation and the matrix form, the voltage equations can be represented by:

$$\begin{bmatrix} u_{sx} \\ u_{sy} \\ u_{rx} \\ u_{ry} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -w_g L_s & pL_m & -w_g L_m \\ w_g L_s & R_s + pL_s & w_g L_m & pL_m \\ pL_m & (P.w_m - w_g)L_m & R_r + pL_r & (P.w_m - w_g)L_r \\ (w_g - P.w_m)L_m & pL_m & (w_g - P.w_m)L_r & R_r + pL_r \end{bmatrix} . \begin{bmatrix} i_{sx} \\ i_{sy} \\ i_{rx} \\ i_{ry} \end{bmatrix} \tag{2.60}$$

2.3.4.2 - Space-phasor voltage equations in the stationary reference frame fixed to the stator.

If $w_g = 0$, from the expression 2.17.

$$\begin{bmatrix} u_{sd} \\ u_{sq} \\ u_{rd} \\ u_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & 0 & pL_m & 0 \\ 0 & R_s + pL_s & 0 & pL_m \\ pL_m & P.w_m L_m & R_r + pL_r & P.w_m L_r \\ -P.w_m L_m & pL_m & -P.w_m L_r & R_r + pL_r \end{bmatrix} . \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \end{bmatrix} \tag{2.61}$$

The stator voltage space phasor can be expressed as follows:

$$\bar{u}_s = R_s \bar{i}_s + \frac{d\overline{\Psi}_s}{dt} \tag{2.62}$$

The rotor voltage space phasor can be written as:

$$\bar{u}'_r e^{-j\theta_m} = R_r \bar{i}'_r e^{-j\theta_m} + \frac{d(\overline{\Psi}'_r e^{-j\theta_m})}{dt} \tag{2.63}$$

$$\bar{u}'_r = R_r \bar{i}'_r + \frac{d\overline{\Psi}'_r}{dt} - jPw_m \overline{\Psi}'_r \tag{2.64}$$

And the flux linkage space phasors can be expressed as follows

$$\overline{\Psi}_s = L_s \bar{i}_s + L_m \bar{i}'_r$$

$$\overline{\Psi}_r = L_r \bar{i}'_r + L_m \bar{i}_s \qquad (2.65)$$

2.4 Torque expressions.

The general expression for the torque is as follows:

$$t_e = c\overline{\Psi}_s \mathbf{x} \bar{i}'_r \qquad (2.66)$$

Where the c is a constant, $\overline{\Psi}_s$ and $\bar{i}'_r$ are the space phasors of the stator flux and rotor current respectively, both referred to the stationary reference frame fixed to the stator.

The expression given above can also be expressed as follows:

$$t_e = c\left|\overline{\Psi}_s\right|\left|\bar{i}'_r\right|\sin\gamma \qquad (2.67)$$

Where γ does the angle exist between the stator flux linkage and the rotor current. It follows that when $\gamma = 90^0$ the torque obtained is the maximum and its expression is exactly equal to the one for the DC machines. Nevertheless, in DC machines the space distribution of both magnitudes is fixed in space, thus producing the maximum torque for all different magnitude values. Furthermore, both magnitudes can be controlled independently or separately.

In an AC machine, however, it is much more difficult to realize this principle because both quantities are coupled and their position in space depends on both the stator and rotor positions. It is a further complication that in squirrel-cage machines, it is not possible to monitor the rotor current, unless the motor is specially prepared for this purpose in a special laboratory. It is impossible to find them in a real application. The search for a simple control scheme similar to the one for DC machines has led to the development of the so-called vector control schemes, where the point of obtaining two different currents, one for controlling the flux and the other one for the rotor current, is achieved.

2.4.1 Deduction of the torque expression by means of energy considerations.

Torque equation is being deduced by means of energy considerations. Therefore, the starting equation is as follows:

$$P_{mech} = P_{electric} - P_{loss} - P_{field} \tag{2.68}$$

Substituting the previous powers for its values, the equation can be expressed as follows:

$$t_e . w_r = \frac{3}{2}\left[\left(\mathrm{Re}\left(\bar{u}_s . \bar{i}_s^{*}\right) - R_s \left|i_s\right|^2 - \mathrm{Re}\left(\frac{d\overline{\Psi}_s}{dt}\bar{i}_s^{*}\right)\right) + \right.$$

$$\left.\left(\mathrm{Re}\left(\bar{u}_r' . \bar{i}_r'^{*}\right) - R_r \left|i_r'\right|^2 - \mathrm{Re}\left(\frac{d\overline{\Psi}_r'}{dt}\bar{i}_s'^{*}\right)\right)\right] \tag{2.69}$$

Since in the stationary reference frame, the stator voltage space phasor u_s can only be balanced by the stator ohmic drop, plus the rate of change of the stator flux linkage, the previous expression can be expressed as follows:

$$t_e . w_r = \frac{3}{2}\mathrm{Re}\left(-jw_r \overline{\Psi}_r' . \bar{i}_r'^{*}\right)$$

$$= \frac{3}{2} w_r \,\mathrm{Re}\left(j\overline{\Psi}_r' . \bar{i}_r'^{*}\right)$$

$$= \frac{3}{2} w_r \overline{\Psi}_r' \mathbf{x} \bar{i}_r' \tag{2.70}$$

Expressing the equation in a general way for any number of pair of poles gives:

$$t_e = -\frac{3}{2} P \overline{\Psi}_r' \mathbf{x} \bar{i}_r' \tag{2.71}$$

If equations 2.63 , 2.64 and 2.35 are substituted in equation 2.71, it is obtained the following expression for the torque:

$$t_e = \frac{3}{2} P \overline{\Psi}_s \mathbf{x} \bar{i}_s \tag{2.72}$$

If the product is developed, expression 2.72 is as follows:

$$t_e = \frac{3}{2} P \left(\Psi_{sD} . i_{sQ} - \Psi_{sQ} . i_{sD}\right) \tag{2.73}$$

Finally, different expressions for the torque can be obtained as follows:

$$t_e = -\frac{3}{2} P \left(L_r \bar{i}'_r + L_m \bar{i}_s \right) \mathbf{x} \bar{i}'_r$$

$$= -\frac{3}{2} P L_m \bar{i}_s \mathbf{x} \bar{i}'_r$$

$$= -\frac{3}{2} P L_m \bar{i}'_s \mathbf{x} \bar{i}_r$$

$$t_e = -\frac{3}{2} P \frac{L_m}{L_s} \left(L_m \bar{i}'_r + L_s \bar{i}_s \right) \mathbf{x} \bar{i}'_r$$

$$= -P \frac{3}{2} \frac{L_m}{L_s} \overline{\Psi}_s \mathbf{x} \bar{i}'_r$$

$$= -\frac{3}{2} P \frac{L_m}{L_s L_r - L_m^2} \overline{\Psi}_s \mathbf{x} \overline{\Psi}_r \tag{2.74}$$

The final torque expression is given by the equation (2.74).

2.5 Conclusion

In the present chapter has been deduced the motor model. The model has been formulated by means of the two-axis theory equations and the space phasor notation. The model is developed in the stator reference frame, and the torque expressions have also been derived.

CHAPTER 3

INDUCTION MOTOR CONTROL

3.1 Introduction

The control and estimation of induction motor drives constitute a vast subject, and the technology has further advanced in recent years. Induction motor dives with cage type machines have been the work horses in the industry for the variable speed applications in a wide power range that covers from fractional horse power to multi megawatts. These applications include pumps and fans, paper and textile mills, subway and locomotive propulsions, electric and hybrid vehicles, machine tools and robotics, home appliances, heat pumps and air conditioners rolling mills, wind generation systems, etc. in addition to process control, the energy saving accept of variable frequency drives is getting a lot of attention now a days.

The control and estimation of ac drives in general is considerable 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, machine parameter variation, and the difficulties of processing feedback signals in the presence of harmonics. While considering drive applications, we need to address the following questions:

- One, two or four quadrant drive?
- Torque, speed, or position control in the primary or outer loop?
- Single or multi motor drive?
- Range of speed control? Does it include zero speed and field-weakening regions?
- Accuracy and response time?
- Robustness with load torque and parameter variations?
- Control with speed sensor and senseless control?

To control the induction thee are different types of control are there they are scalar control, vector control, direct torque control and flux control.

3.2 Induction Motor controls:

The general control block diagram for variable frequency speed control of an induction motor drive is shown below. It consists of converter machine system with hierarchy of control loops added to it. The converter machine is shown with voltage (V^*_s) and frequency (ω_e^*) as control inputs. The outputs are shown as speed (ω_r), developed torque (T_e), stator current (I_s) and rotor flux (Ψ_r). Instead of voltage control loop, the converter may be current controlled with direct and indirect voltage control in the inner loop [12]-[13].

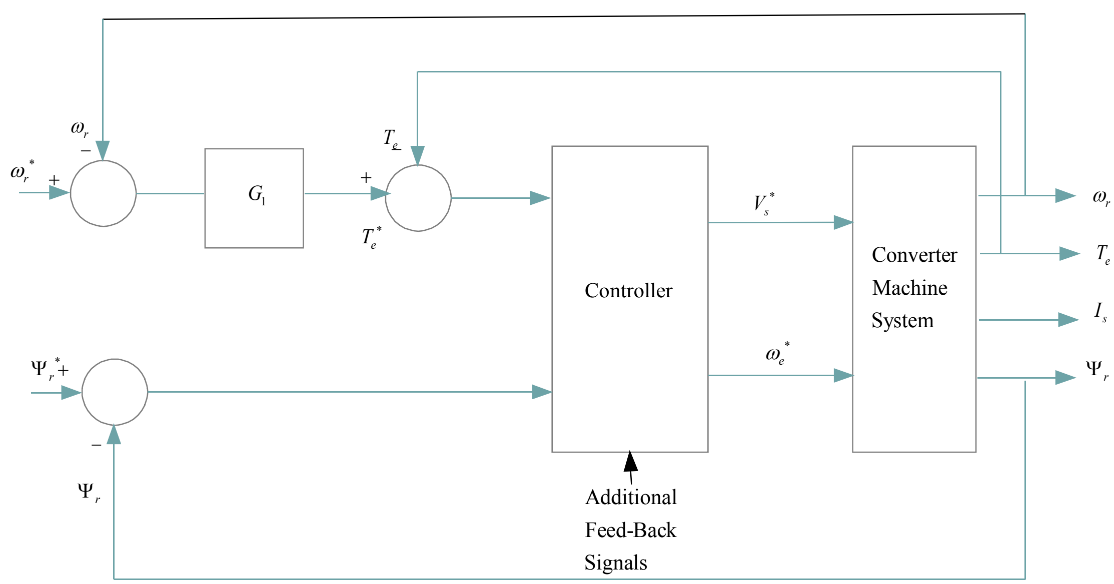


Figure 3.1 General speed control block diagram of Induction motor

The speed control is shown with an inner torque control loop, which may be optional. Adding a higher gain inner loop control provides the advantage of linearization, improved bandwidth and the ability to control the signals with in safe limits, like a dc machine, the flux of an ac machine is normally controlled to be constant at a rated value because it gives the faster response and high developed torque per ampere of current. In fact the flux under consideration may be stator flux (Ψ_s), rotor flux (Ψ_r), or the air gap flux (Ψ_m or Ψ_g). How ever, the rotor flux control is considered in the present case. the inner control loops have faster response (i.e., higher bandwidth) than the outer loop. Since an ac drive system is an multi variable, nonlinear with internal coupling effect, and discrete time in nature, its stability analysis is very complex.

3.2.1 Voltage Control.

There are many different ways of control of induction motor. But the basic control is the stator voltage control of the induction motor. By this the speed can be varied but confined to a very small stable region. The main disadvantage with this model is, by varying voltage the flux will change which tends to the inefficient control of the induction motor [9]-[11].

3.2.2 Voltage/frequency.

There are many different ways to drive an induction motor. The main differences between them are the motor's performance and the viability and cost in its real implementation. Despite the fact that "Voltage/frequency" (V/f) is the simplest controller, it is the most widespread, being in the majority of the industrial applications. It is known as a scalar control and acts imposing a constant relation between voltage and frequency.
The structure is very simple and it is normally used without speed feedback. However, this controller doesn't achieve a good accuracy in both speed and torque responses mainly due to the fact that the stator flux and the torque are not directly controlled. Even though, as long as the parameters are identified, the accuracy in the speed can be 2% (except in a very low speed) and the dynamic response can be approximately around 50ms [4]-[8] .

3.2.2 Field Acceleration method.

This method is based on avoiding the electromagnetic transients in the stator currents, keeping its phase continuous. Therefore, the equations used can be simplified saving the vector transformation in the controllers.

It is achieved some computational reduction, overcoming the main problem in the vector controllers and then becoming an important alternative for the vector controllers [17].

3.3 Vector Control

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

The foregoing problems can be solved by vector control or field – oriented control. The invention of vector control is in the beginning of 1970s, and the demonstration that the induction motor can be controlled like a separately excited dc motor, brought a renaissance in high performance control of ac drives. Because of dc machine like performance, vector control is also known as decoupling, orthogonal or transvector control. Vector control is applicable to both induction and synchronous motor drives. Undoubtedly, vector control and the corresponding feedback signal processing, particularly for modern sensor less vector control, are complex and the use of powerful microcomputer or DSP is mandatory. It appears that eventually, vector control will oust scalar control, and will be accepted as industry standard control for ac drives.

3.3.1 DC Drive Analogy

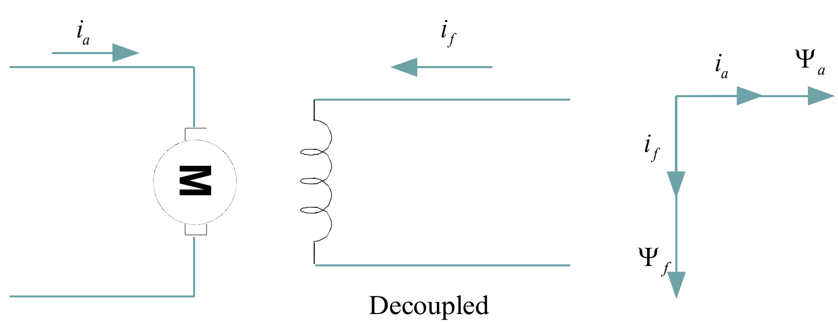
Ideally, a vector – controlled induction motor drive operates like a separately excited dc motor drive, as mentioned above. Figure shown below explains this analogy. In dc machine, neglecting the armature reaction effect and field saturation, developed torque is given by

$$T = K_e I_a I_f \tag{3.1}$$

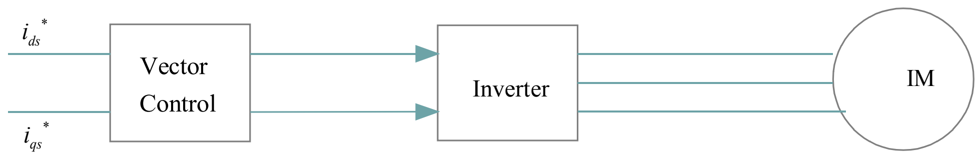
were I_a is the armature current and I_f is the field current. The construction of a dc machine is such that the field flux ψ_f produced by the current I_f is perpendicular to the flux ψ_a, which is produced by the armature current I_a. These space vectors, which are stationary in space, are orthogonal or decoupled in nature. this means that when torque is controlled by controlling the current I_a, the flux ψ_f is not effected and we get the fast transient response and high torque/ampere ratio with the rated ψ_f.

Because of decoupling, when the field current I_f is controlled, it effects the flux ψ_f only but not the ψ_a flux. Because of the inherent coupling problem, an induction motor cannot generally give such fast response.

DC machine like performance can also be extended to an induction motor if the machine control is considered in the synchronously rotating reference frame ($d^e - q^e$), where the sinusoidal variables appear in as dc quantities in steady state. The induction motor with the inverter and vector control in the front end is shown with two control inputs, i_{ds}^* and i_{qs}^*.



(a)



(b)

Figure. 3.2 (a) Separately excited DC motor, (b) Vector-controlled induction motor

These currents are the direct axis components and quadrature axis component of the stator current, respectively, in a synchronously rotating reference frame. With vector control. i_{ds} is analogous to field current i_a of the dc machine. Therefore the torque can be expressed as

$$T_e = K_t \psi_r i_{qs} \tag{3.2}$$

or

$$T_e = K'_t i_{ds} i_{qs} \tag{3.3}$$

Where ψ_r is absolute value and ψ'_r is the peak value of the sinusoidal space vector. This dc machine like performance is only possible if i_{ds} is oriented (or aligned) in the direction of the flux ψ_r and i_{qs} is established perpendicular to it, as shown in the above figure. This means that when i_{qs}^* is controlled; it affects the actual i_{qs} current only, but does not affect the flux ψ_r. Similarly, when $i_{ds}*$ is controlled, it controls the flux only and does not affect the i_{qs} component of current.

This vector of field orientation of currents is essential under all operating conditions in a vector - controlled drive. Note that when compared to dc machine space vector orientation, induction machine space vectors rotate synchronously at frequency ω_e, as indicated in the figure. In summery, vector control should assure the correct orientation and equality of command and actual currents.

.

The developed torque may be expressed as

$$T_{em} = k_a \phi(I_f) I_a \tag{3.4}$$

Where k_a is the constant coefficient, $\phi(I_f)$ the field flux, and I_a, the armature current. Here, the torque angle is naturally 90degrees, flux may be controlled by adjusting the field current, I_f and torque can be controlled independently of flux by adjusting the armature current, I_a since the time constant of the armature circuit is much smaller than that of field winding, controlling torque by changing armature current is quicker than changing I_f, or both.

3.3.2 Principles of Vector Control

The fundamentals of the vector control implementation can be explained with the help of fig 3.4 where the machine model is represented in synchronously rotating reference frame. The inverter is omitted from the figure assuming the unity current gain, that is it generates i_a, i_b, and i_c as dictated by the corresponding command currents i_a^*, i_b^*, and i_c^* from the controller. A machine model with internal conversion is shown in the right. The machine terminal phase currents i_a, i_b, and i_c are converted to i_{ds}^s and i_{qs}^s components by 3phase / 2phase transformation. These are then converted to synchronously rotating frame by the unit vector Components $\cos\theta_e$ and $\sin\theta_e$ before

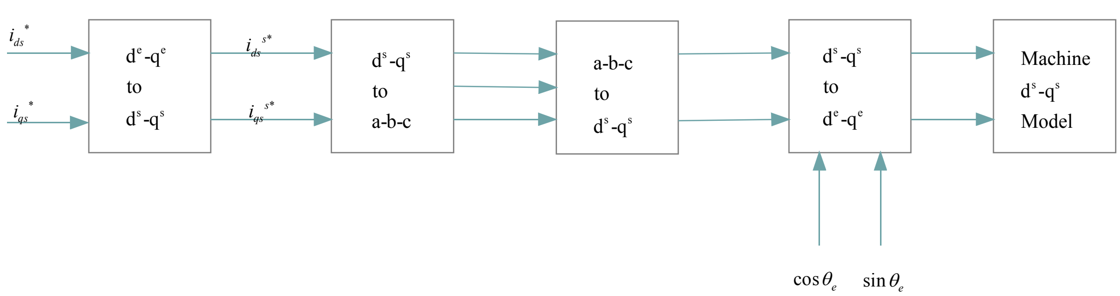


Figure 3.3 Vector control implementation principle with machine $d^e - q^e$ model

applying them to the $d^e - q^e$ machine model as shown. The controller makes two stages of inverse transformation, as shown. So that the control currents i_{ds}^* and i_{qs}^* correspond to the machine currents i_{ds} and i_{qs}, respectively. In addition unit vector assures correct alignment of i_{ds} current from the flux vector $\overline{\Psi}_r$ and i_{qs} perpendicular to it. The response of i_{ds} and i_{qs} is instantaneous.

There are essential two general methods of vector control. One called the direct or feer-back method, was invented by Blaschke, and the other known as indirect or feed forward method, was invented by Hasse. The methods are different essentially by how the vector ($\cos\theta_e$ and $\sin\theta_e$) is generated for the control. It should be mentioned here that orientation gives natural decoupling control, where as air gap or stator flux orientation gives a coupling effect which has to be compensated by decoupling compensation current.

3.3.3 Indirect or Feed Forward Vector Control

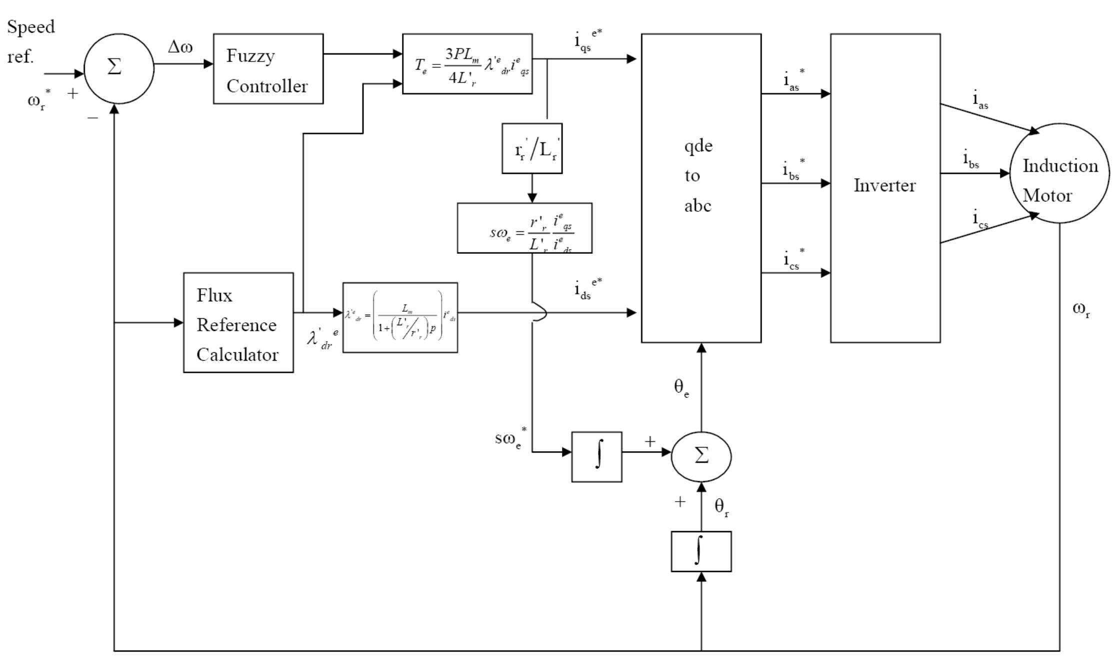


Figure 3.4 shows indirect vector control of induction motor drive

Since the rotor flux depends on the stator current and the machine speed, the spatial orientation between the rotor and stator fields is not 90 degrees by default in an induction motor. Field oriented control aims at maintaining the orthogonal spatial angle between the rotor and the stator fields, and achieving independent control of the rotor flux. Such requirements are satisfied by controlling the magnitude and phase of the applied stator current.

Transforming the 3-phase variables of a symmetrical induction machine in a q-d axis reference frame rotating synchronously with the rotor flux, the electromagnetic torque developed by the machine can be written as:

$$T_e = \frac{3PL_m}{4L'_r}\left(\lambda_{dr}'^{e}\, i^{e}_{qs} - \lambda_{qr}'^{e}\, i^{e}_{ds}\right) \tag{3.5}$$

Where P is the number of poles, L_m is the magnetizing inductance, λ'^{e}_{dr}, λ'^{e}_{qr}, i^{e}_{ds} and i^{e}_{qs} are the rotor flux and stator current direct (d) and quadrature (q) components, respectively, and

$$L'_r = L_m + L'_{lr} \tag{3.6}$$

L'_{lr} being the rotor equivalent leakage inductance. The rotor flux components can be expressed as:

$$\lambda'^{e}_{qr} = L'_r\, i'^{e}_{qr} + L_m i^{e}_{qs} \tag{3.7}$$

and

$$\lambda'^{e}_{dr} = L'_r\, i'^{e}_{dr} + L_m i^{e}_{ds} \tag{3.8}$$

Where i'^{e}_{dr} and i'^{e}_{qr} are the d-q components of the rotor current.

In a field oriented induction machine, the synchronously rotating reference frame is chosen so that its d-axis is aligned with the rotor flux. Hence λ'^{e}_{qr} is zero, and from Equations (3.5) and (3.7),

$$T_e = \frac{3PL_m}{4L'_r}\lambda'^{e}_{dr}\, i^{e}_{qs} \tag{3.9}$$

And $$i'^{e}_{qr} = -\frac{L_m}{L'_r} i^{e}_{qs} \tag{3.10}$$

The corresponding positions of the stator current and rotor flux are shown in Fig 3.5.

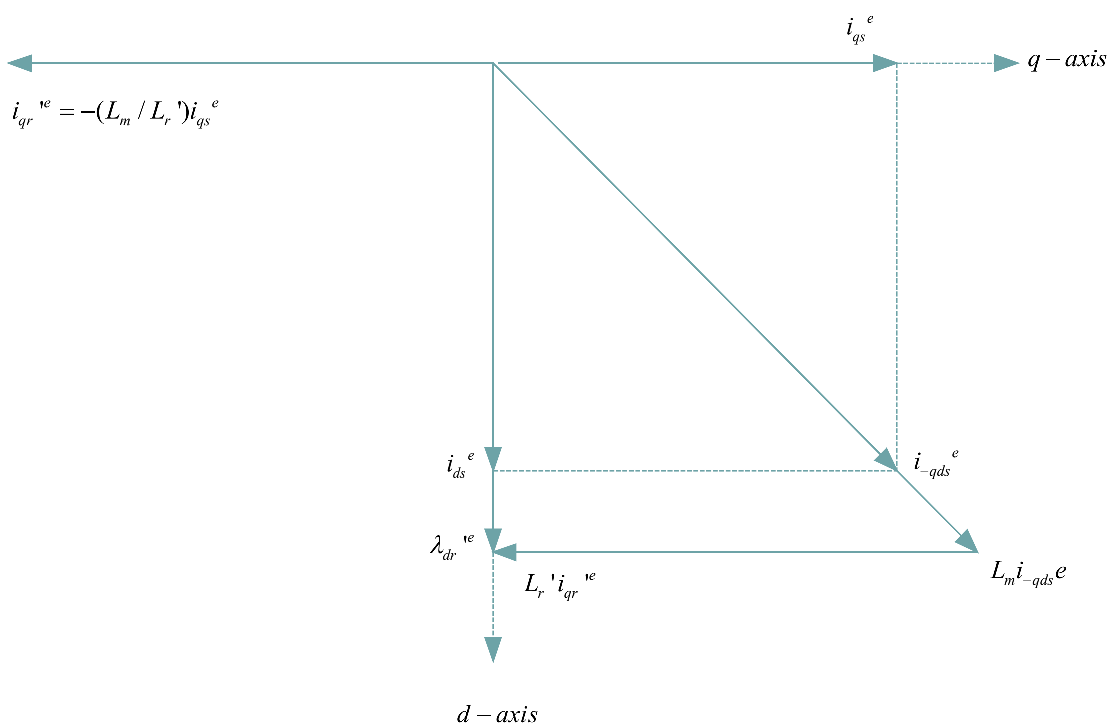


Fig 3.5: Motor currents and flux in synchronously rotating reference frame

The rotor induced voltage is zero for an induction machine with short circuited rotor bars. In the synchronously rotating reference frame, the q-axis and d-axis components of the rotor voltage can therefore be written as

$$v'^{e}_{qr} = 0 = r'_{r}\, i'^{e}_{qr} + \frac{d\lambda'^{e}_{qr}}{dt} + s\omega_{e}\lambda'^{e}_{dr} \tag{3.11}$$

$$v'^{e}_{dr} = 0 = r'_{r}\, i'^{e}_{dr} + \frac{d\lambda'^{e}_{dr}}{dt} + s\omega_{e}\lambda'^{e}_{qr} \tag{3.12}$$

Where $s\omega_e$ is slip frequency. The d-axis component of the rotor flux is obtained from Equations (3.8) and (3.12) as

$$\lambda'^{e}_{dr} = \left(\frac{L_m}{1 + \left(L'_r / r'_r \right) p} \right) i^{e}_{ds} \tag{3.13}$$

Where $p = \frac{d}{dt}$. With $\lambda'^{e}_{qr} = 0$, the steady state rotor flux is therefore

$$\lambda'^{e}_{dr} = L_m i^{e}_{ds} \tag{3.14}$$

And is aligned with the d-axis rotor current, as shown in the Fig 3.5. Hence, orthogonality between the rotor and stator fields can be achieved.

Equation (3.9) shows that the developed torque is proportional to quadrature component of the stator current and rotor flux. Equation (3.14) also shows that the rotor flux can be independently controlled by i^{e}_{ds}, as required for field orientation. Replacing i'^{e}_{qr} by equation (3.10) and λ'^{e}_{dr} by equation (3.14), the slip frequency for proper field orientation can be expressed as:

$$s\omega_e = \frac{r'_r}{L'_r}\frac{i^e_{qs}}{i^e_{ds}} \tag{3.15}$$

From the knowledge of the rotor speed, ω_r, the required reference frame angle, θ_{rf} for field oriented can be found as

$$\theta_{rf} = \int_0^t \left(s\omega_e + \omega_r\right) dt \tag{3.16}$$

A block diagram of proposed Indirect Field oriented control is shown in the Figure 3.4. speed information, obtained by tachogenerator feedback, enables computation of torque reference using a fuzzy controller. The later is designed to cater for non linear characteristics of the machine and estimations in the motor parameters used in the computation blocks. The drive essentially uses feed forward control ((3.9) to (3.16)) to apply the correct value of the reference frame angle so as to control the torque output and rotor flux of the machine.

The d-q current components are then transformed into 3-phase command currents (i^*_a, i^*_b and i^*_c) which are applied to the current regulated pulse width modulated (CRPWM) dc to ac inverter. The later in turn applies the correct values of stator currents to the motor for proper field orientation.

3.3.4 Direct Torque Control.

In Direct Torque Control it is possible to control directly the stator flux and the torque by
selecting the appropriate inverter state.
Its main features are as follows:

- Direct torque control and direct stator flux control.
- Indirect control of stator currents and voltages.

- Approximately sinusoidal stator fluxes and stator currents.
- High dynamic performance even at locked rotor.

This method presents the following advantages:

- Absence of co-ordinate transform.
- Absence of voltage modulator block, as well as other controllers such as PID for flux and torque.
- Minimal torque response time, even better than the vector controllers.
  Although, some disadvantages are present:
- Possible problems during starting.
- Requirement of torque and flux estimators, implying the consequent parameters identification.
- Inherent torque and flux ripples.

3.2.3.1 Introduction As it has been introduced in expression 2.72, the electromagnetic torque in the three phase induction machines can be expressed as follows [21][24]:

$$t_e = \frac{3}{2} P \overline{\Psi}_s \mathrm{x} \bar{i}_s \tag{3.19}$$

Where $\overline{\Psi}_s$ is the stator flux, $\bar{i}_s$ is the stator current (both fixed to the stationary reference frame fixed to the stator) and P the number of pairs of poles. The previous equation can be modified and expressed as follows:

$$t_e = \frac{3}{2} P \left|\overline{\Psi}_s\right| . \left|\bar{i}_s\right| . \sin(\alpha_s - \rho_s) \tag{3.20}$$

Where ρ_s is the stator flux angle and α_s is the stator current one, both referred to the horizontal axis of the stationary frame fixed to the stator.

If the stator flux modulus is kept constant and the angle ρ_s is changed quickly, then the electromagnetic torque is directly controlled.

The same conclusion can be obtained using another expression for the electromagnetic torque. From equation 2.74, next equation can be written:

$$t_e = \frac{3}{2} P \frac{L_m}{L_s L_r - L_m{}^2} \left| \overline{\Psi}_s \right| \mathbf{x} \left| \overline{\Psi}_r{}' \right| . \sin(\rho_s - \rho_r) \tag{3.21}$$

Because of the rotor time constant is larger than the stator one, the rotor flux changes slowly compared to the stator flux; in fact, the rotor flux can be assumed constant. (The fact that the rotor flux can be assumed constant is true as long as the response time of the control is much faster than the rotor time constant). As long as the stator flux modulus is kept constant, then the electromagnetic torque can be rapidly changed and controlled by means of changing the angle $\rho_s - \rho_r$ [24].

3.4 Conclusion:

In this chapter the types of control of induction motors are being explained. The vector control of induction motor i.e. indirect vector control of induction motor is dealt in depth and the brief introduction of direct torque control of induction motor is explained.

CHAPTER 4

THEORY OF FUZZY LOGIC

4.1 Introduction

Although a large number of artificial intelligence techniques have been employed in modern power systems, fuzzy logic is a powerful tool in meeting challenging problems in modern power systems. This is so because fuzzy logic is the only technique that can handle imprecise, vague of fuzzy information. In the past decade fuzzy system have supplemented conventional technologies in many scientific applications and engineering systems, especially in control systems and pattern recognition. The same fuzzy technology in the form of approximate reasoning is also resurfacing in information technology, where it provides decision support and expert systems with powerful reasoning capabilities bound by a minimum number of rules. Various concepts relating to fuzzy logic were discussed briefly in this chapter. This heuristic technique has its own pros and cons, which are also mentioned in this chapter.

4.2 What Is Fuzzy Logic?

"Fuzzy logic is a mathematical discipline, based on fuzzy set theory, which allows for degrees of truth or falsehood". As opposed to "binary logic" which holds that an assertion must be either true or false. 'Fuzzy logic' accommodates the possibility that the assertion can be partly false. The degree of truth or falsehood in the assertion can be both qualitatively and quantitatively described. As mentioned above fuzzy logic deals with uncertainty, ambiguity and imprecision, which widely exist in complex real world problems especially in engineering. Fuzzy logic is a multi-valued logic primarily concerned with the uncertainty and approximate reasoning. The intelligent soft technique mimic's human decision making process and is biologically inspired.

4.2.1 Fuzzy Logic as an Aspect of Uncertainty

Fuzziness is often confused with probability. The fundamental difference between them is that fuzziness deals with deterministic plausibility, while probability concerns with the likelihood of non-deterministic stochastic events. The uncertainty of probability generally relates to the occurrence of phenomena, as symbolized by the concept of randomness. Fuzziness and randomness differ in nature i.e. they are different aspects of uncertainty. The former conveys "subjective" human thinking, feeling, or language and the latter indicates an "objective" statistic in the natural science. Fuzziness describes event's ambiguity. It measures the degree to which an event occurs is 'random'. To what degree is occurs is 'FUZZY'.

4.2.2 Set Theory of Fuzzy Logic

Fuzzy logic set theory is a theory of classes of objects with unsharp boundaries fuzzy set, introduced by L.A.Zadeh (father of fuzzy logic) in the year 1965 as a mathematical way to represent vagueness in linguistics can be considered a generalization of classical set theory. A fuzzy set is a generalization of an ordinary set in that it allows the degree membership for each element to range over the unit interval [0,1].

A classical (crisp) set is a collection of distinct objects. It is defined in such a way as to dichotomize the elements of a given universe of discourse: members and nonmembers. Finally a crisp set can be defined by the so called characteristic function μ_a (x) of a crisp set in U takes its values in {0,1} and is defined such that $\mu_a(x)=1$ if x is a member of A(i.e. x belongs to A) and 0 other wise. That is

$$\mu_A(x) = \begin{cases} 1, \text{ if and only if } X \in A \\ 0, \text{ if and only if } X \in A \end{cases}$$

Note that:

[a] The boundary of set A is rigid and sharp and performs a two class dichotomization (i.e., X€ A or X does not belong to A), and

[b] The universe of discourse U is a crisp set.

A fuzzy set, on the other hand, introduces vagueness by eliminating the sharp boundary that divides members from nonmembers in the group. Thus, the transition between full memberships is gradual rather than abrupt. Hence, fuzzy sets may be viewed as an extension and generalization of the basic concepts of crisp ~~s~~ets. A fuzzy set A in the universe of discourse U can be defined as a set of ordered pairs,

A={X, μ_A(x) ||X€U}.

Where μ_A( ) is called the membership function (or characteristic function) of A and μ_AA(X) is the grade (or degree) of the membership of ~~X~~ in A which indicates the degree that X belongs to A. The membership function μ_A( ), maps U to the membership space M that is $\mu_A : \mu \rightarrow M$. when M={0,1}, set A is non-fuzzy and μ_A( ) is the characteristic function of the crisp set A. the distinction between crisp and fuzzy sets is given by the below example. Let U be the real line R and let crisp set A represent "real numbers greater than and equal to 5", then we have A={X, μ_A(x) ||X€U},

Where the characteristic function is

$$\mu_A(x) = \begin{bmatrix} 0, X < 5 \\ 1, X > 5 \end{bmatrix}$$

Which is shown in fig4.1 (a). Now let fuzzy set A represent "real numbers close to 5". Then we have

A={X, μ_A(x) ||X€U},

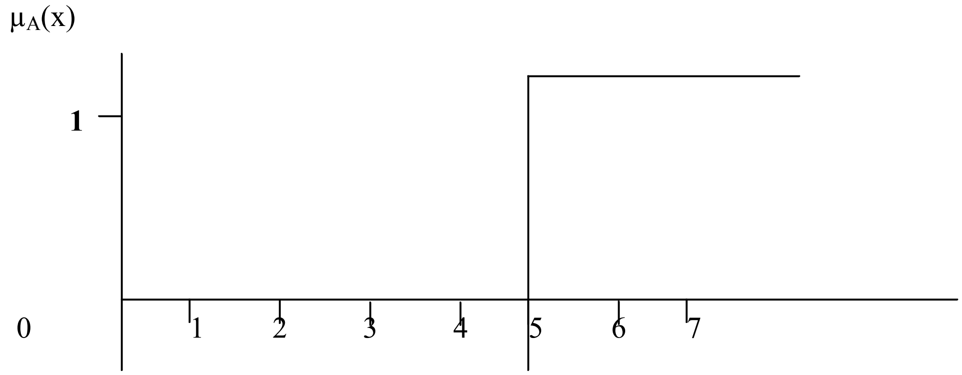


Figure 4.1(a) Characteristic functions of crisp set A

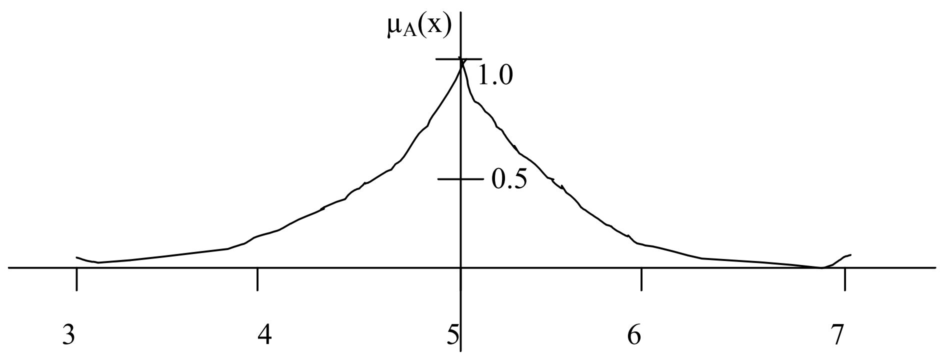


Figure4.1 (b) Characteristic functions of fuzzy set A

$\mu_A(x)=1/(1+10(X-5)^2)$,

The fuzzy set A in the above example can also be represented as

$A=\{X, \mu_A(x) \|X€\}=\{ [(1+10(X-5)^2]^{-1}\}$

We can use another membership function for the fuzzy set A:

For example,

$$\mu_A(x)=1/(1+(X-5)^2).$$

These two different membership functions show that assignment of the membership function of a fuzzy set is subjective in nature, however, it cannot be assigned arbitrarily. A qualitative estimation reflecting a given ordering of the elements in A may be sufficient. Further more, estimated membership functions are complicated, and a better approach is to utilize the learning power of neutral networks to approximate them.

In engineering applications of fuzzy logic, the most commonly used shapes for membership functions are triangular, trapezoidal, piecewise linear Gaussian and sinusoidal or bell shaped.

4.2.3 Mathematical Operations on Fuzzy Sets

With the basic notations and definitions for fuzzy sets some fundamental operation as fuzzy sets are given below. The classical set operations such as union, intersection and compliment are applicable to fuzzy set theory as well, which are often referred as t-operator, t-conform and negation. Let us assume A and B to be two fuzzy sets defined in some universe of discourse (UOD). The various mathematical operations on the above fuzzy sets A and B are as follows:

1.Equality: A and B are equal if and only if.

$\mu_A(x)= \mu_B(x)$ for all $X \in U$.

2. Subset: A is a subset of B i.e., A belongs to B if and only if

$\mu_A(x) <= \mu_B(x)$ for all $X \in U$.

3. Complement: When $\mu_A(x) = [0, 1]$, the complement of A denoted as A\` is defined by membership functions.

$$\mu_A(x) = 1-\mu_B(x) \text{ for all } X \in U.$$

4. Union: The union of fuzzy sets A and B, denotes as A union B is defined as

$$\begin{aligned}\mu_{AUB}(x) &= \max [\mu_A(x), \mu_B(x)] \\ &= \mu_A(x) \vee \mu_B(x) \text{ for all } X \in U.\end{aligned}$$

Where v indicates max operation.

5. Intersection: The intersection of fuzzy sets A and B, denoted as A intersection B is defined as

$$\mu_{A\cap B}(x)=\min[\mu_A(x), \mu_B(x)]$$

$$= \mu_A(x) \wedge \mu_B(x) \text{ for all } X\in U.$$

Where $\wedge$ indicates min operation. It is clear that

6. Double negation law (involution):

$$\overline{A}=A, \overline{B}=B.$$

7.De Morgan's laws:

a) $\overline{A\cup B}=\overline{A}\cap\overline{B}$

b) $\overline{A\cap B}=\overline{A}\cup\overline{B}$

8. By centriod method:

$$A=\frac{\sum_{i=1}^{n} X_i \mu_A(X_i)}{\sum_{i=1}^{n} \mu_A(X_i)}$$

where A is the fuzzy set defined in

X={x_1, x2, x3 …xn}

A still simpler method assumes that

$\mu_A(a_i)=\max \mu_A(X)$ where $X_i \in X$.

Law of the excluded middle ($E\cup E=U$) and the law of contradiction ($E\cap E=\varnothing$) of the crisp set E are no longer true and valid in fuzzy sets i.e., for fuzzy set A,

$A\cup\overline{A}=U$ ~~and~~ $A\cap\overline{A}=\varnothing$

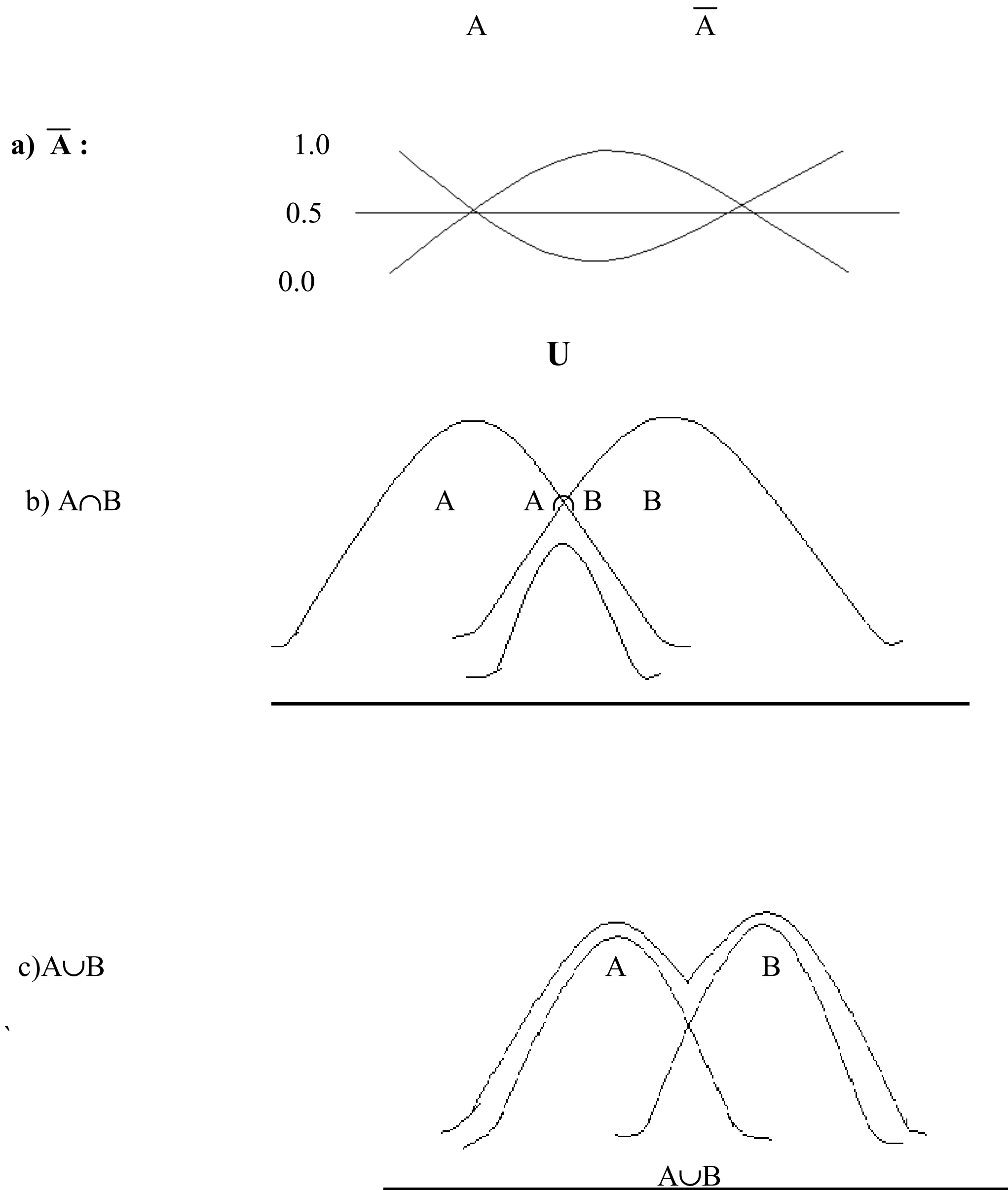


Figure4.2: Complement, intersection and union of two fuzzy sets A and B.

4.2.4 Fuzzy Modeling

Fuzzy logic system is widely used in fuzzy logic controllers and signal processing applications. A fuzzy logic system maps into crisp inputs and crisp outputs. Hence there are 8 major tasks typically needed for developing fuzzy logic systems.

- Define the problem.

- Define the linguistic variables.
- Define control surfaces (fuzzy sets).
- Define behavior of the control surfaces (fuzzy rules).
- Define reasoning mechanism (fuzzy inference).
- Build the system.
- Test the system.
- Tune and validate the system.

A fuzzy logic system has four major components:

1. Rules of knowledge base.
2. Fuzzifier.
3. Inference engine
4. Defuzzifier.

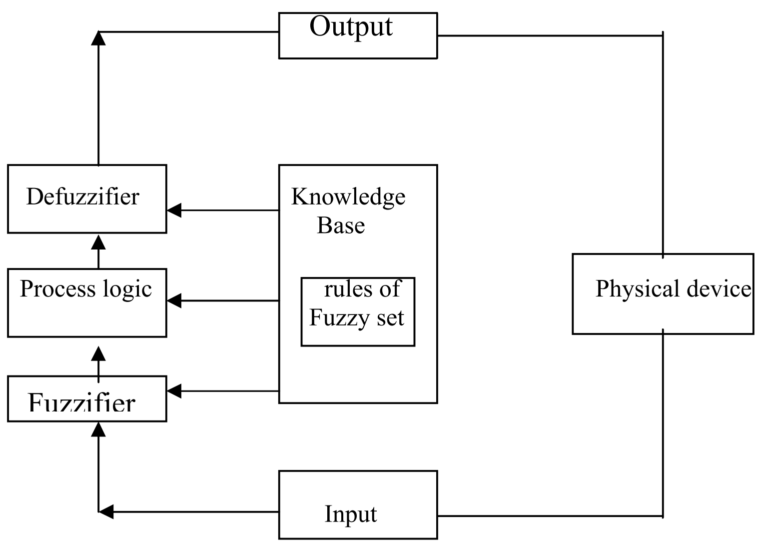


Figure 4.3 Typical Fuzzy system

Fuzzifier:

Fuzzifier transforms real to fuzzy domain associating a membership function to each fuzzy set through a mapping F: $R \in I^n$ where n is number of fuzzy sets associated with a given quantity $X \in R$ to the interval $I=[0,1]$. The Fuzzification interface involves the following functions:

1. Measurement of the values of input functions, variables
2. Performing a scale mapping the transfer range of values of input variables corresponding universe of discourse and
3. Performing the function if Fuzzification that converts input data into suitable linguistic values, which may be viewed as labels of fuzzy sets.

Knowledge base:

This encodes the experts' knowledge in the form of a set of **if-then** rules. Fuzzy logic systems store rules as fuzzy association; i.e., for the rule **if A then B,** where A and B are fuzzy sets, a fuzzy logic system stores the association (A, B) in a matrix M, the fuzzy associate matrix M maps fuzzy set A and fuzzy set B. This association of fuzzy rules is called a Fuzzy Association Memory **(FAM).**

Inference engine:

Fuzzy inference is Kernel in a fuzzy logic system. It has capability of simulating human decision-making based on fuzzy concepts and of interring fuzzy control actions employing fuzzy implication and rules of inference in fuzzy logic. In the fuzzy inference engine, fuzzy logic principles are used to combine fuzzy if-then rules from the fuzzy rule-base into mapping from fuzzy input sets to fuzzy output sets.General methods for inferences are given are max-min inference, max-product inference.

Defuzzification:

This maps fuzzy domain to real domain, which may be thought as an inverse Fuzzification. Defuzzification can be described as a mapping of vector X_f with n sets to real number given by

$DF: I^n \rightarrow R$

Defuzzification is an art rather than a science. One criterion for the choice of a defuzzifier is computational simplicity. A Defuzzification strategy is aimed at producing a non-fuzzy control action that best represents the possibility distribution of an interred fuzzy control action. There are three popular methods for defuzzification.

Viz.,

1.Mean of maximum (MOM).

2. Max criterion &

3.Center of area (COA)

Amongst these the center of area (COA) is most widely used strategy, which generates the center of gravity of the possibility distribution of a control action.

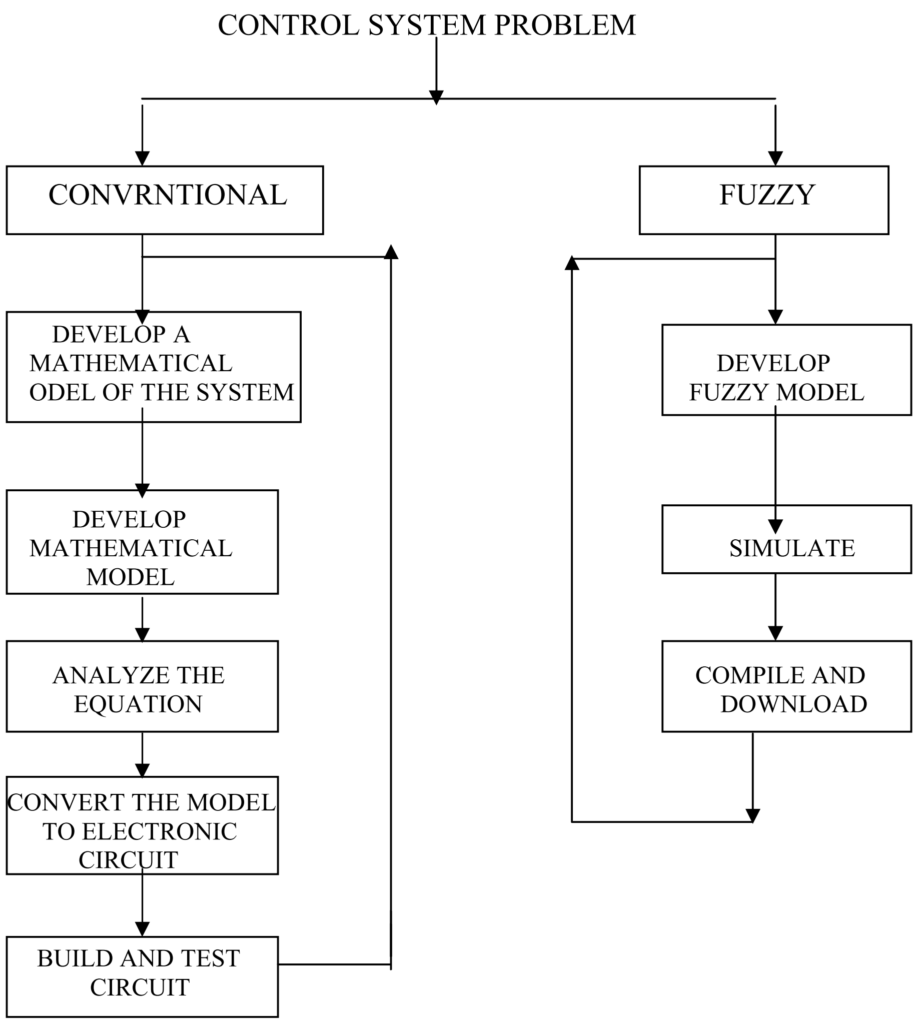


Figure 4.4 Block diagram: Comparison of a conventional system with a fuzzy logic one

Design of fuzzy logic system involves understanding of the process dynamics, choice of universe of discourse (UOD) with number of partitions, defining membership function of the fuzzy sets, selection of appropriate

Fuzzification and Defuzzification process, choice of scale factors, the correct formulation of rule-base i.e., deriving if then linguistic rules to obtain better performances.

The advantages of fuzzy logic include quicker development cycles and superior performance. Typically, conventional control systems have five development steps:

Developing a model of the system, developing a mathematical model for the controller, analyzing the equations, converting the model to a circuit and building a circuit. Fuzzy logic on other hand has just three: developing a fuzzy model, stimulating it and compiling it.

4.2.5 Applicability Of Fuzzy Logic

One major feature of fuzzy logic is its ability to express the amount of ambiguity in human thinking and subjectivity (including natural language) in a comparatively undistorted manner. Thus, it is appropriate to use fuzzy logic:

[1] When the process is concerned with continuous [phenomena (e.g., one or more of the control variables are continuous) that are not easily broken down into discrete segments].

[2] When a mathematical model of the process does not exist or exists but is too difficult to encode or is too complex to be evaluated fast enough for real time operation or involves too much memory on the designed chip architecture.

[3] When high ambient noise levels be dealt with or it is important to use in expensive sensors and/or log-precision microcontrollers.

When the process involves human interaction (e.g., human descriptive or intuitive thinking)

[4] When an expert is available who can specify the rules underlying the system behavior and the fuzzy sets that represent the characteristic of each variable.

4.2.6 Practical Applications of Fuzzy Logic

Commonsense, human thinking and judgment are the lures of fuzzy logic, fuzzy logic techniques applications in such areas as:

- ❖ Control (the most widely applied areas)
- ❖ Pattern recognition (e.g., image, audio, signal processing)
- ❖ Quantitative analysis (operation, research, management)
- ❖ Inference (e.g., expert systems for diagnosis, planning and prediction; natural language processing, intelligent interface, intelligent robots, software engineering)
- ❖ Information retrieval (e.g., databases)

The advantages of fuzzy logic are:

- ► Fuzzy logic is based on natural languages and is conceptually easy to understand.
- ► Fuzzy logic is tolerated of imprecise data and can handle ambiguity.
- ► Fuzzy logic can be built on top of the experience of experts or can be implemented with other techniques.
- ► Fuzzy logic can resolve conflicting objectives.
- ► Fuzzy logic us flexible and is relatively easy to implement.

4.3 Fuzzy Speed Controller

The motor-control issues are traditionally handled by fixed-gain proportional-integral (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. The problem can be solved by several adaptive control techniques such as model reference adaptive control (MRAC) [54], sliding-mode control (SMC) [55], variable structure control (VSC) [56], and self-tuning PI controllers [57], etc. The design of all of the above controllers depends on the exact system mathematical model. However, it is often difficult to develop an accurate system mathematical model due to unknown load variation, unknown and unavoidable parameter variations due to saturation, temperature variations, and system disturbances. In order to overcome the above problems, recently, the fuzzy-logic controller (FLC) is being used for motor control purpose [7]–[12]. The mathematical tool for the FLC is the fuzzy set theory introduced by Zadeh [58]. 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 [59]. The earlier reported works for fuzzy-logic applications in motor drives [60]–[63] are mainly theoretical and based on either simulation or experimental results at low-speed operating conditions.

The proposed fuzzy speed controller is shown in Figure 3. The error, $\Delta\omega$, between the set point and shaft speed, is used to generate the following inputs to the controller:

$$u_1 = K_P \Delta\omega \tag{4.1}$$

$$u_2 = K_I \int \Delta\omega dt \tag{4.2}$$

By proper adjustment of the member functions and the gains K_P and K_I the controller realizes a non linear PI control algorithm. The output is a torque reference, T^*_{em}, for meeting the desired speed tracking action of the drive.

The input membership functions of the fuzzy system are triangular, with increasing widths away from the origin, so as to obtain finer control action near the set point. In the fuzzification process, these membership functions crisp inputs, u1and u2, into fuzzy subsets with the following linguistic variables: Positive Big (PB), Positive Medium(PM), Positive Small (PS), Zero (ZE), Negative Small (NS), Negative Medium (NM) and Negative Big (NB). The knowledge base consists of 49 rules for determining the result of the fuzzy implication process, as shown in the Table 1. For instance, the first entry in the rule base reads

Where

Positive Big (PB) =9,
Positive Medium (PM) =6,
Positive Small (PS) =3,
Zero (ZE) =0,
Negative Small (NS) =-3,
Negative Medium (NM) =-6
Negative Big (NB) =-9
IF u1is “NB” AND u2 is “NB” then u3 is “NB”.

Since a finite crisp output is required, the output membership function are designed to follow a uniform distribution without end point saturation. In this paper, the *Min* operator is used for computing the premise and implication results, while the *Max* operator is used for aggregation process. The resulting overall fuzzy set is transformed into a real torque command value using Center of Gravity (CoG) defuzzification, which provides adequate compromise between accuracy and computational effort [26]. The rules are shown in the fig 4.6 by using the fuzzy tool box The input output mapping of the fuzzy controller is shown in the figure 4.7. the controller surface is almost linear for small disturbances about zero error.

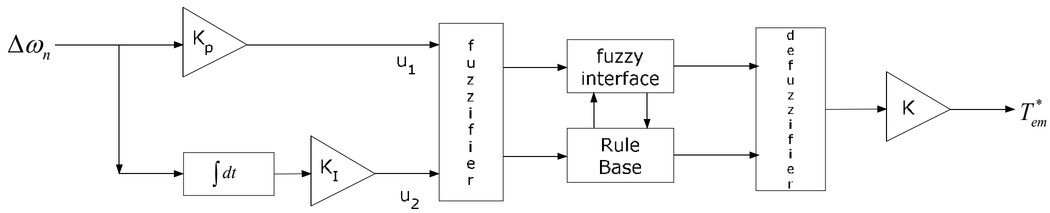


Figure 4.5 Fuzzy Controller Block diagram

Rule Base

u1\u2	NB	NM	NS	ZE	PS	PM	PB
NB	NB	NB	NB	NB	NM	NS	ZE
NM	NB	NB	NB	NM	NS	ZE	PS
NS	NB	NB	NM	NS	ZE	PS	PM
ZE	NB	NM	NS	ZE	PS	PM	PB
PS	NM	NS	ZE	PS	PM	PB	PB
PM	NS	ZE	PS	PM	PB	PB	PB
PB	ZE	PS	PM	PB	PB	PB	PB

Rules:

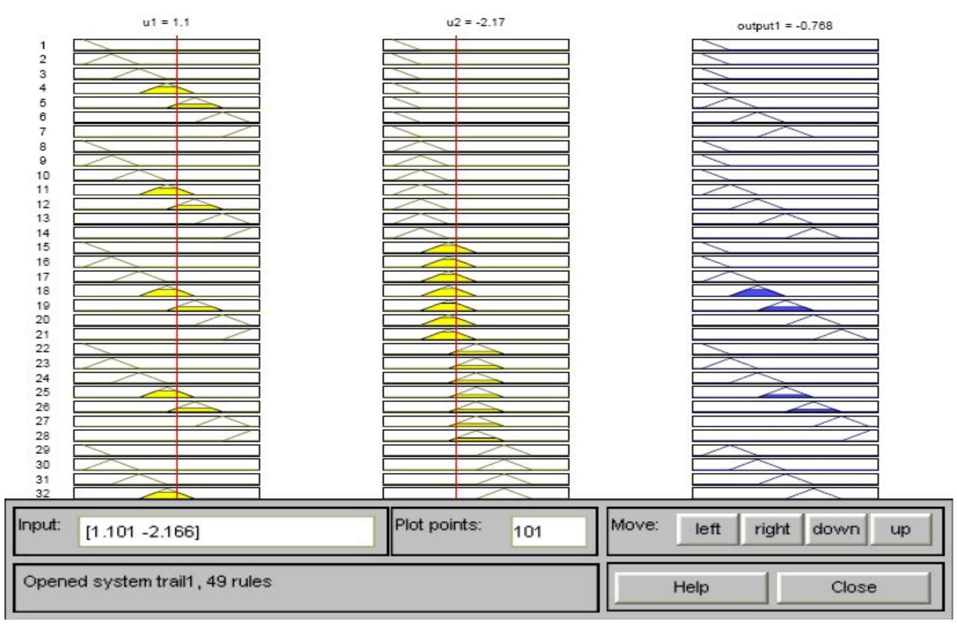


Figure 4.6 Fuzzy rules

Surface Plot

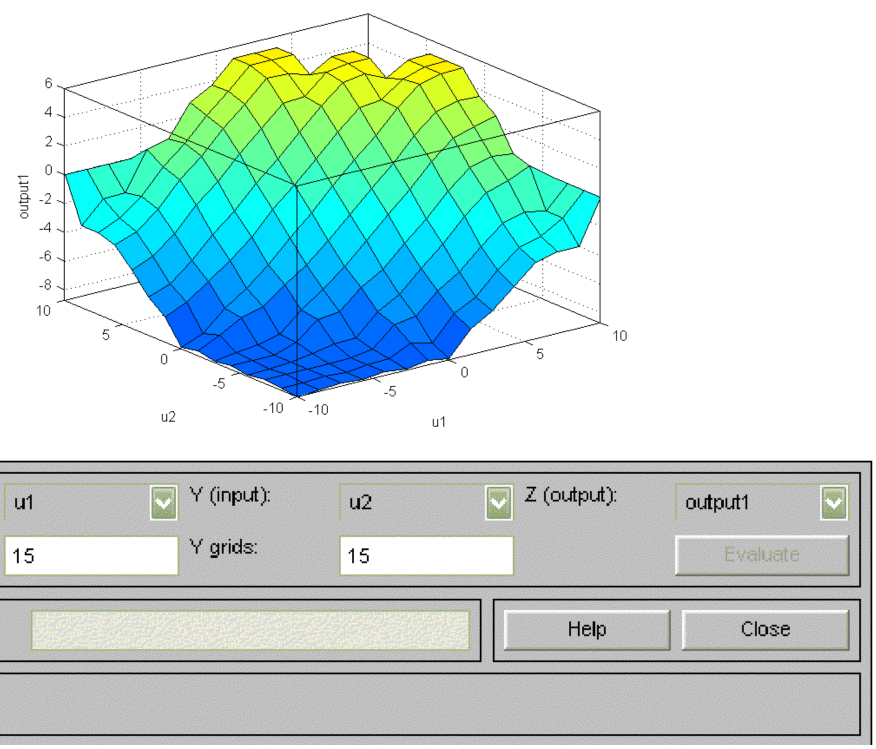


Figure 4.7 surface plot for the rule base

4.4 Conclusion:

In this chapter the complete Artificial Intelligence technique fuzzy logic is discussed. i.e. What is Fuzzy logic and how is it is different from other conventional methods, applicability of fuzzy logic in engineering applications and then the application of fuzzy logic controller how it is designed , rules used.

CHAPTER 5

SIMULINK MODEL

5.1 Description of Simulink Model

The Fuzzy logic PI (proportional Integral) controller torque controller converts the speed error to a reference torque, T^{*}_{em}. Going into the field orientation block are the reference torque, T^{*}_{em}, the d – axis rotor flux, λ'^{e}_{dr}, and the rotor angle, θ_r.

$$i^{s*}_{qs} = i^{e*}_{qs} \cos\rho + i^{e*}_{ds} \sin\rho$$

$$i^{s*}_{ds} = -i^{e*}_{qs} \sin\rho + i^{e*}_{ds} \cos\rho \tag{5.1}$$

$$i^{e}_{qs} = i^{s}_{qs} \cos\rho - i^{s}_{ds} \sin\rho$$

$$i^{e}_{ds} = i^{s}_{qs} \sin\rho + i^{s}_{ds} \cos\rho \tag{5.2}$$

The figure given below shows the inside view of the Field – Oriented block. Inside it, Eqs. 3.15, 3.16, and 3.17 are used to compute the values of i^{e*}_{ds}, i^{e*}_{qs}, and ω_2^{*}. The angle, ρ, is the sum of the slip angle, θ_2, and the rotor angle θ_r. In the qde2abc block, the transformations given in the Eqs. 5.1 And 5.2 are used to generate the abc reference currents.

In the simulation the three large shunt resistors, each of a value that is 500 times the base impedance of the machine, are connected across the stator phase terminals to the stator point, g, of the source. They are used to generate the input terminal abc phase voltages to the stator windings of the induction motor.

The look-up table for field – weakening matches the desired value of the rotor d – axis flux, λ'^{e}_{dr}, to that of the mechanical speed of the rotor, ω_{rm}. For speeds less than the base or the rated speed, λ'^{e}_{dr} is set equal to its no – load value with rated supply voltage. Beyond the base speed, the flux speed product is held constant at the base speed value.

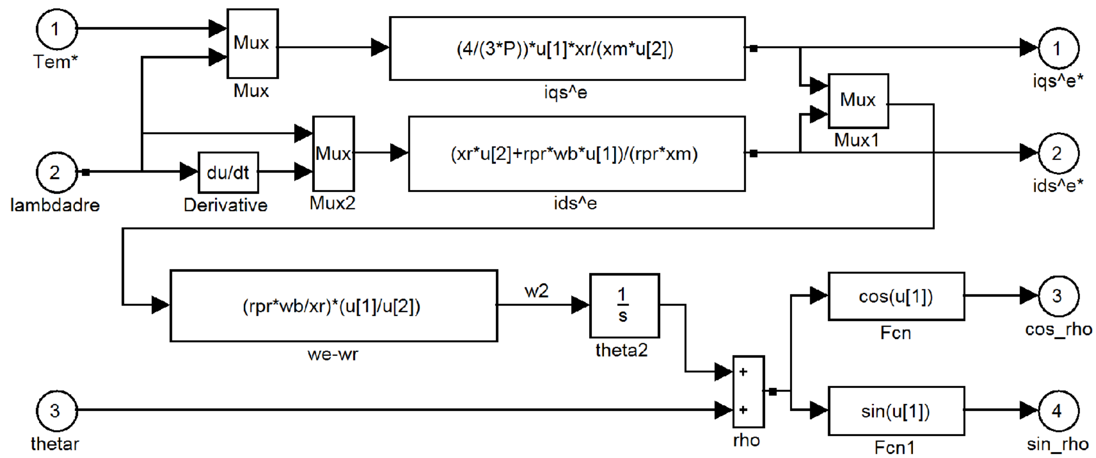
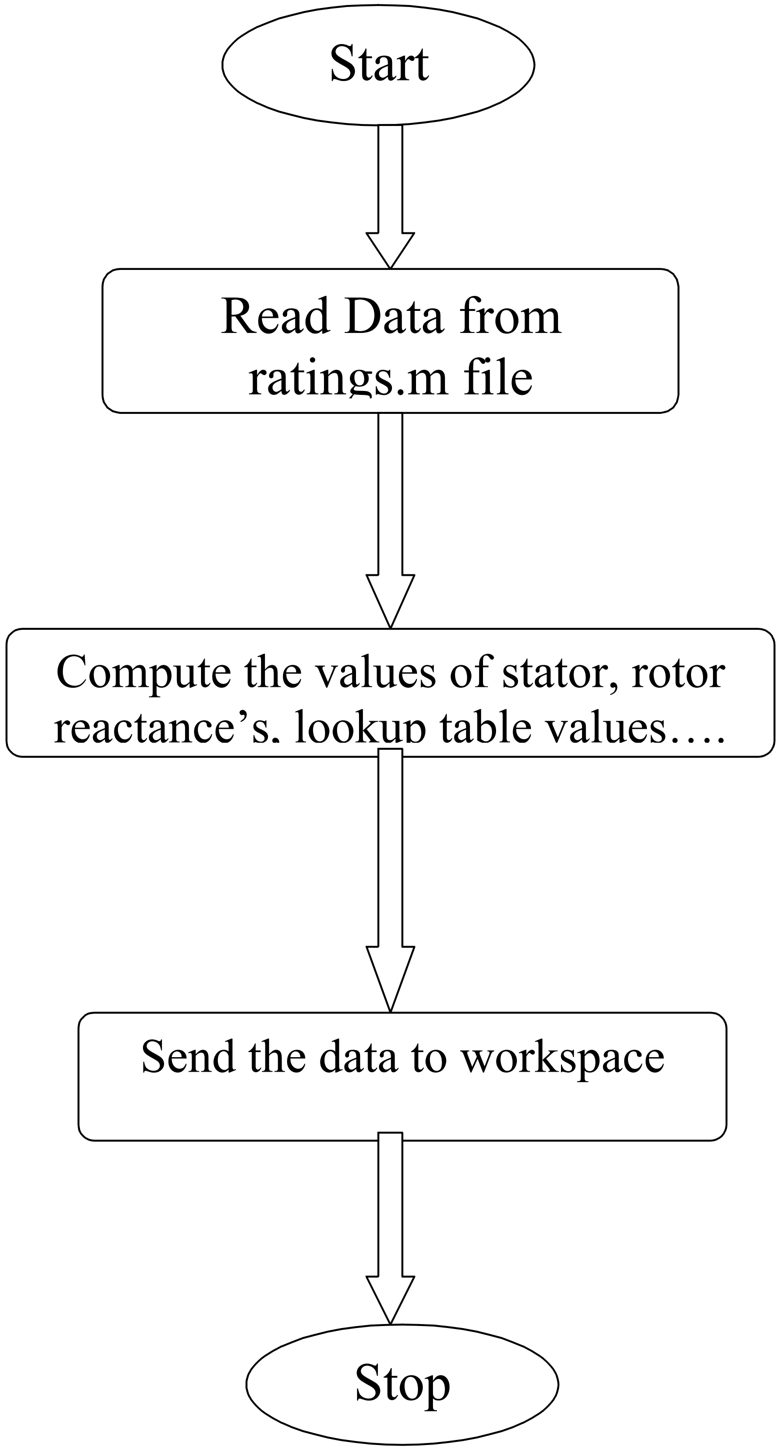


Figure 5.1 Field oriented block

The value of λ'^{e}_{dr} and mechanical speed are generated by another m – file, which is also used to setup the parameters for the simulation in workspace.

5.2Flow Chart For Parameter Calculation



5.3 Algorithm For Parameter Calculation:

STEP1: Start

STEP2: Read the data from the *ratings.m* file

STEP3: Calculate the values of the reactances of stator and rotor, power…

STEP4: Send the data to the workspace

STEP5: Stop

Conclusion:

The construction of simulation model (simulink model) is described and the flowchart , algorithm for the calculation of the parameters such as reactance's, reference torque, values in the lookup table are calculated.

CHAPTER 6

RESULTS AND DISCUSSION

6.1 Introduction

In the previous chapters the modeling of the induction motor, modeling of vector control block and the fuzzy controller is done. And the simulation is preceded as follows. The gains K_P and K_I of the fuzzy controller were adjusted iteratively until the desired speed tracking performance were obtained.

The parameters used in the simulation of the proposed drive as follows;

Rated line to line voltage: 220V, 50Hz

Power rating: 15kVA (4pole)

Stator referred Resistances: r_s =0.10 Ω ; $r_s' = 0.08\,\Omega$

Stator referred leakage inductance: $L_s = L_r' = 725\mu H$

Magnetizing inductance L_m =18.6mH

Rotor inertia, J=2.6 kgm^2

Before the simulation started make sure that the ratings.m , start.m, fuzzy.fis and simu.mdl are in the same directory then start the simulation by excecuting the file start.m then open the fuzzy toolbox excute the fuzzy.fis file then export the data to the workspace. Then run the simulation of the file simu.mdl.

CASE 1:

STEP CHANGES IN TORQUE AT A FIXED REFERENCE SPEED.

To investigate the load torque rejection capability of the drive, the speed is first ramped from 0 to 157 rad/s, as shown in the Figure 5. At t=0.75sec., a step load torque corresponding to the rated value is applied to the rotor shaft. The load on the rotor shaft is next reduced to 50% at the time t = 1sec. in both the cases, the motor torque response is instantaneous and the speed is restored back to the reference value in minimal time. the transient response during the start up is shown in the figure 5.1. This response is very much reduced

Speed conditions

time_wref:	[0 0.5 tstop]
speed_wref:	[0 wbm wbm]

Torque conditions

time_tmech:	[0 0.75 0.75 1 1 1.25 1.25 1.5 1.5 2]
tmech_tmech:	[0 0 –Trated –Trated –Trated/2 –Trated/2 –Trated –Trated 0 0]

CASE 2:

CYCLIC CHANGE OF SPEED REFERENCE.

In the second case Figure 5.6. The reference speed is ramped up to 79rad/sec in the forward motoring mode, followed by linear deceleration to zero, a ramp in the opposite direction and finally reverse braking to standstill. In all four modes, the motor speed is seen to follow the reference with negligible error.

Speed conditions

time_wref1:	[0 0.25 0.5 1.0 1.25 1.5]
speed_wref1:	[0 wbm/2 wbm/2 -wbm/2 -wbm/2 0]

Torque conditions

time_tmech1:	[0 tstop]
tmech_tmech1:	[0 0]

6.2 Results

CASE 1:

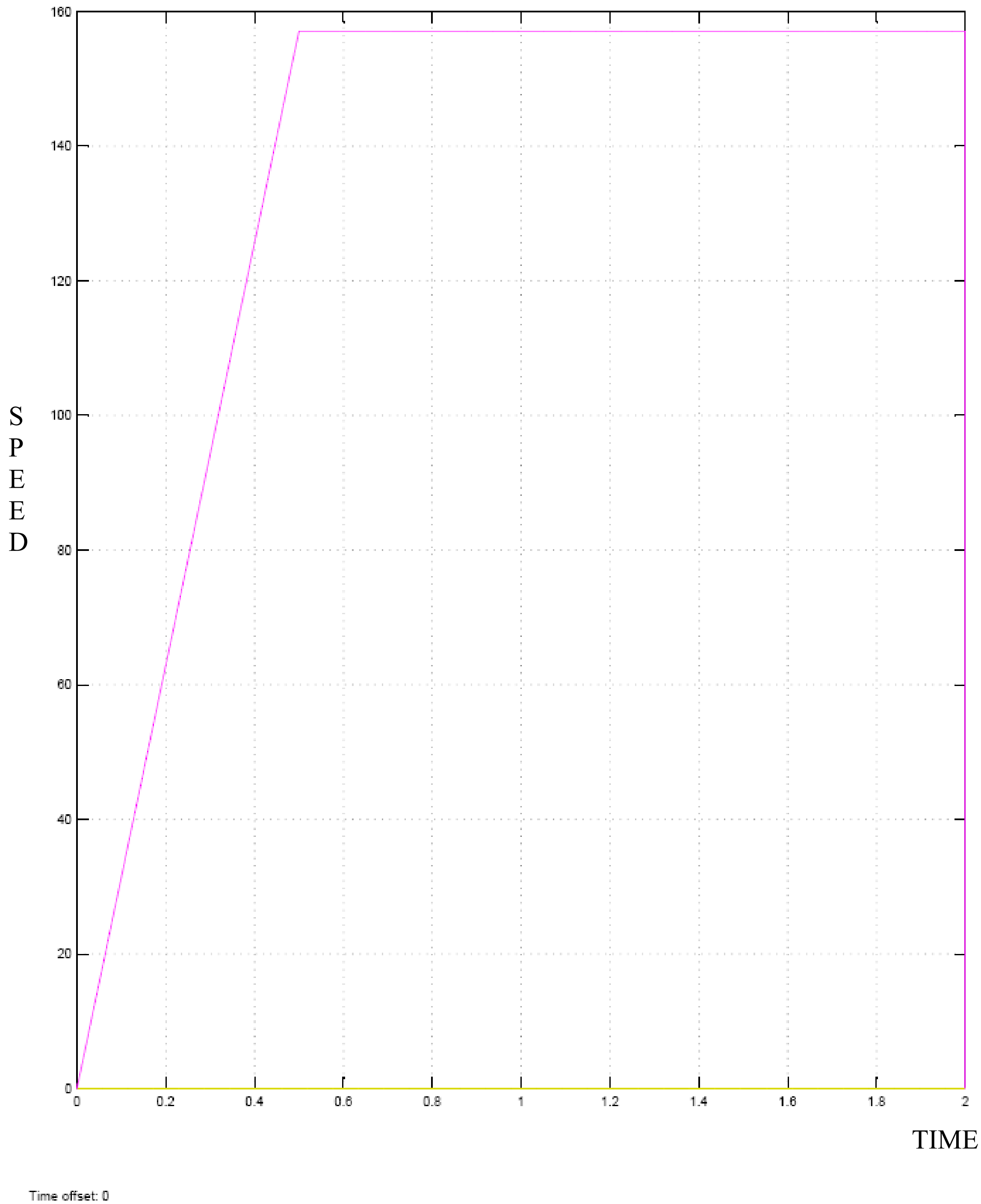


Figure 6.1 reference speed of the motor

The above figure shows that the speed is ramped up to 158rad/sec in 0.5 sec and the const speed is set to maintain

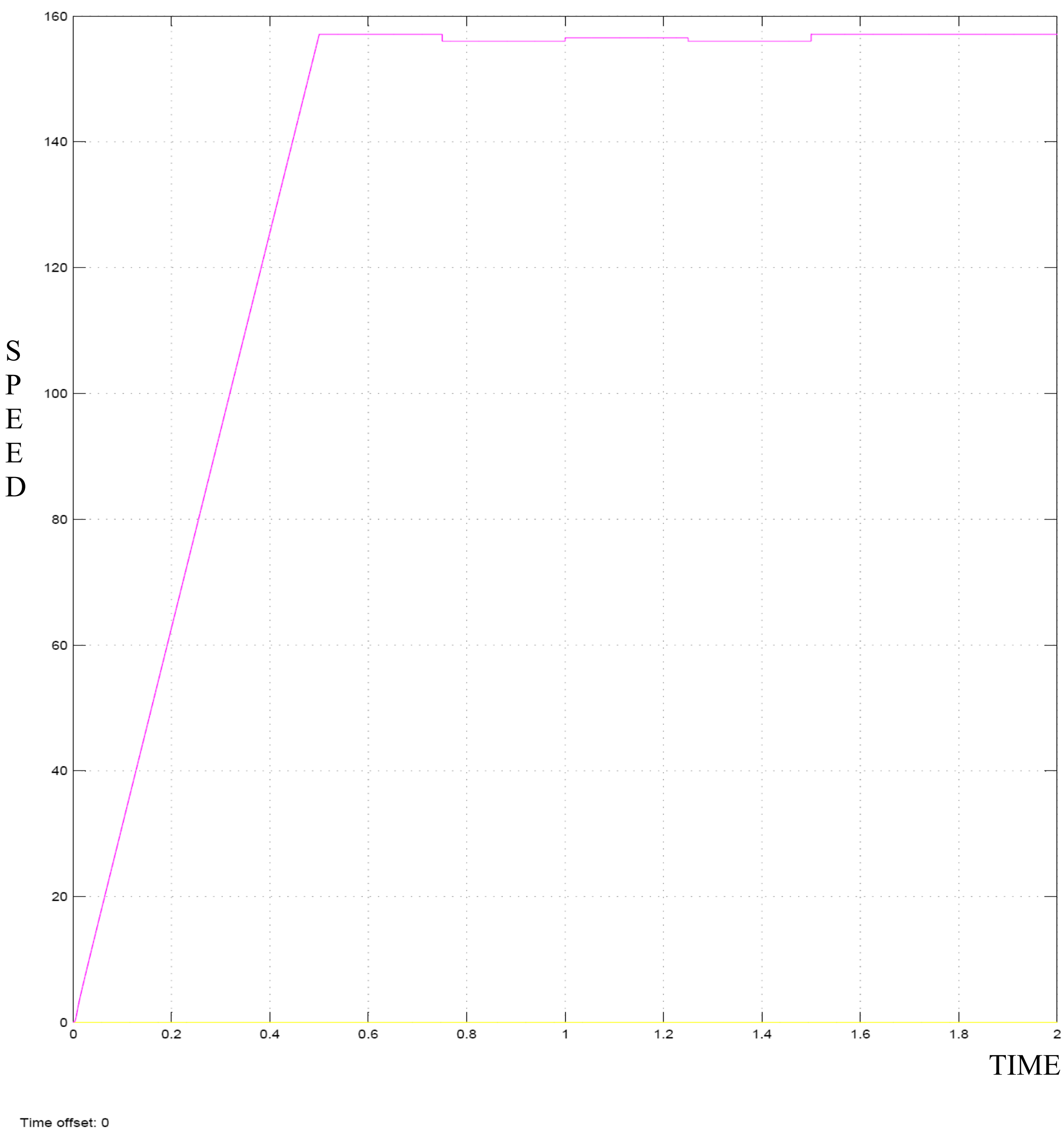


Figure 6.2 speed of the motor

The figure shows the output speed of the motor of the motor. Till the full load is appied the motor followed the reference speed and the after application of tha full load there is a very small dip in the full load speed.

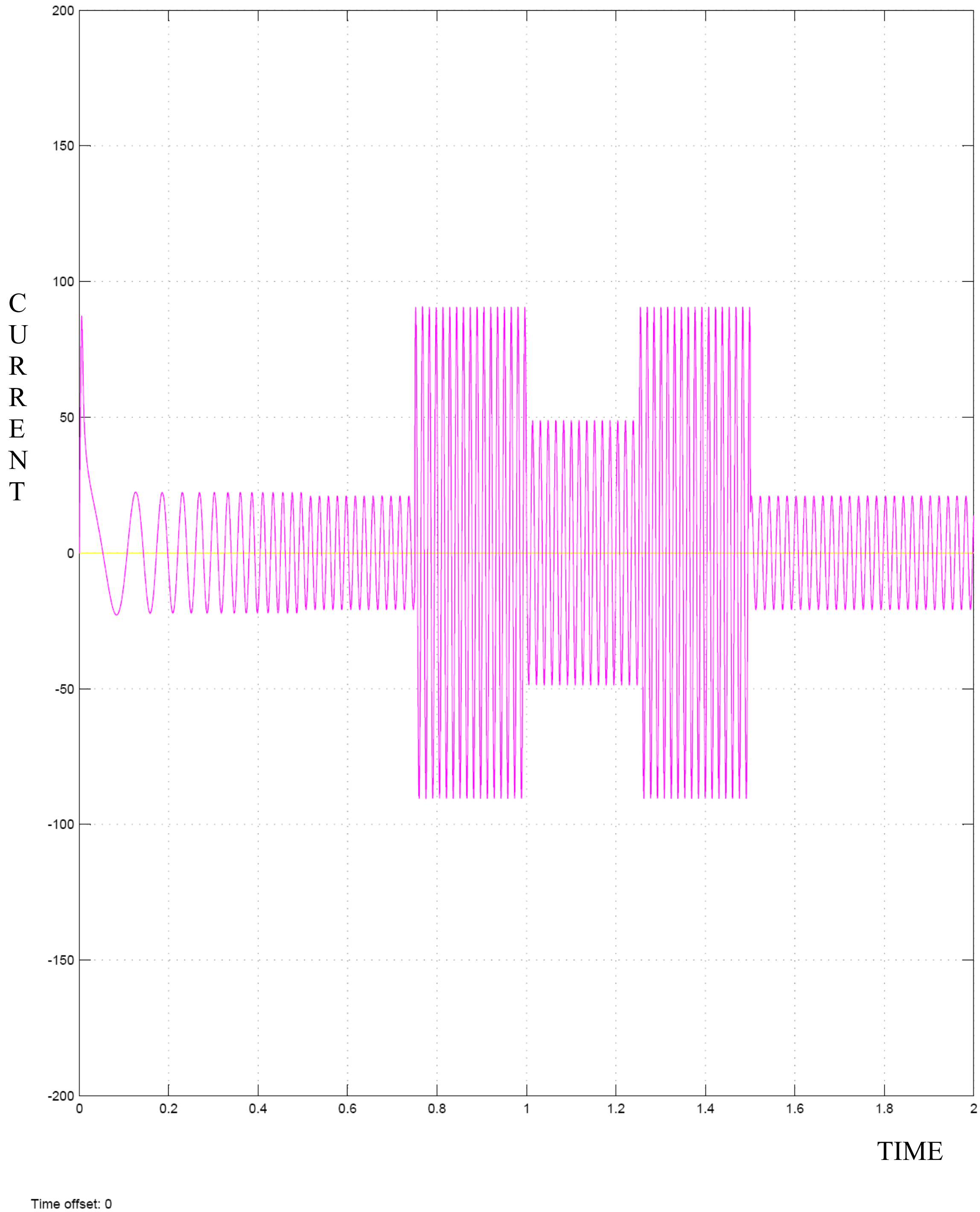


Figure 6.3 stator current

During the starting the current neared the full load current. Then it followed the loading of the motor

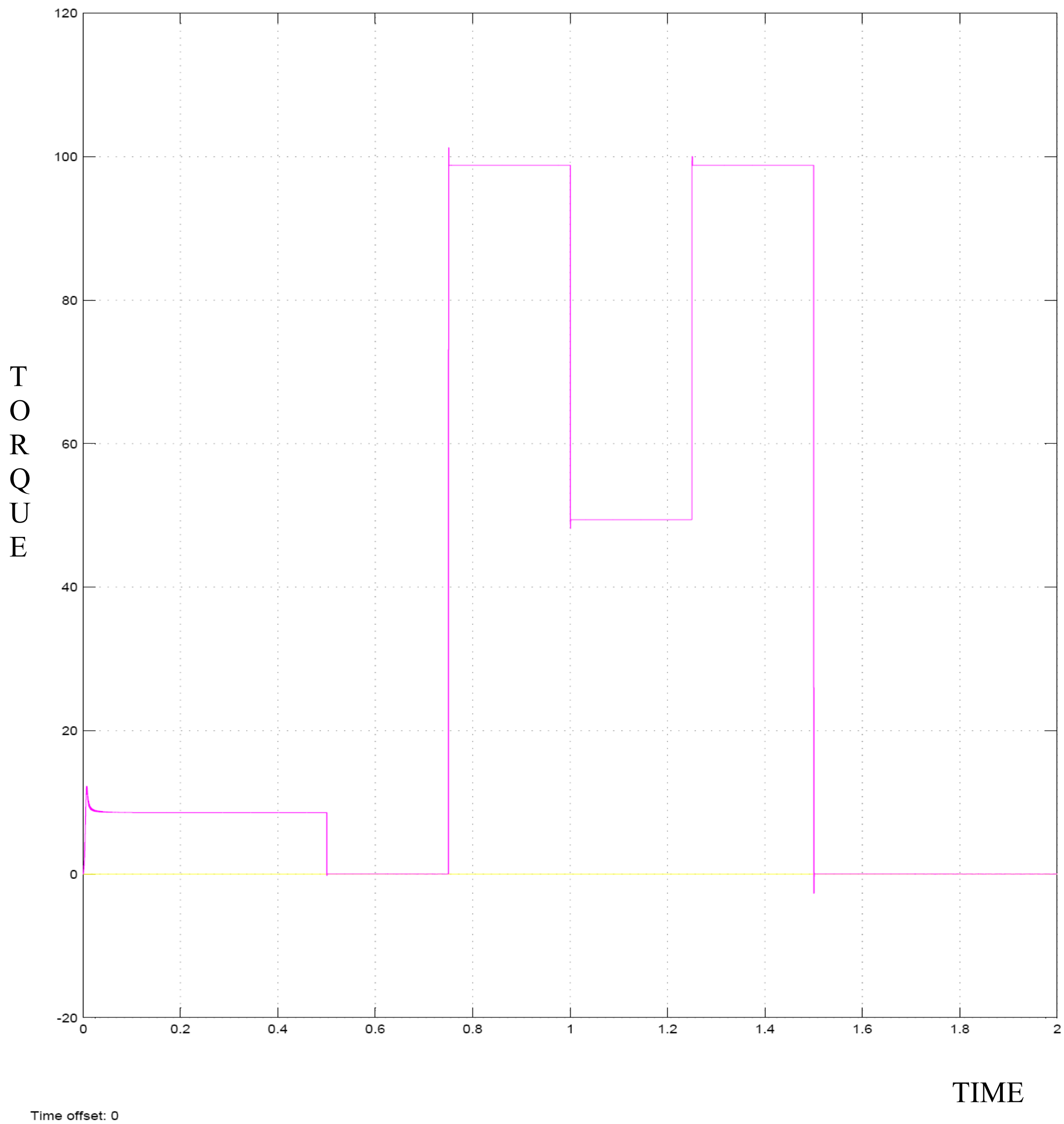


Figure 6.4 motor torque

During the starting the transient torque is very less after that it followed the step changes in the torque with an excellent dynamic response and with zero steady state error

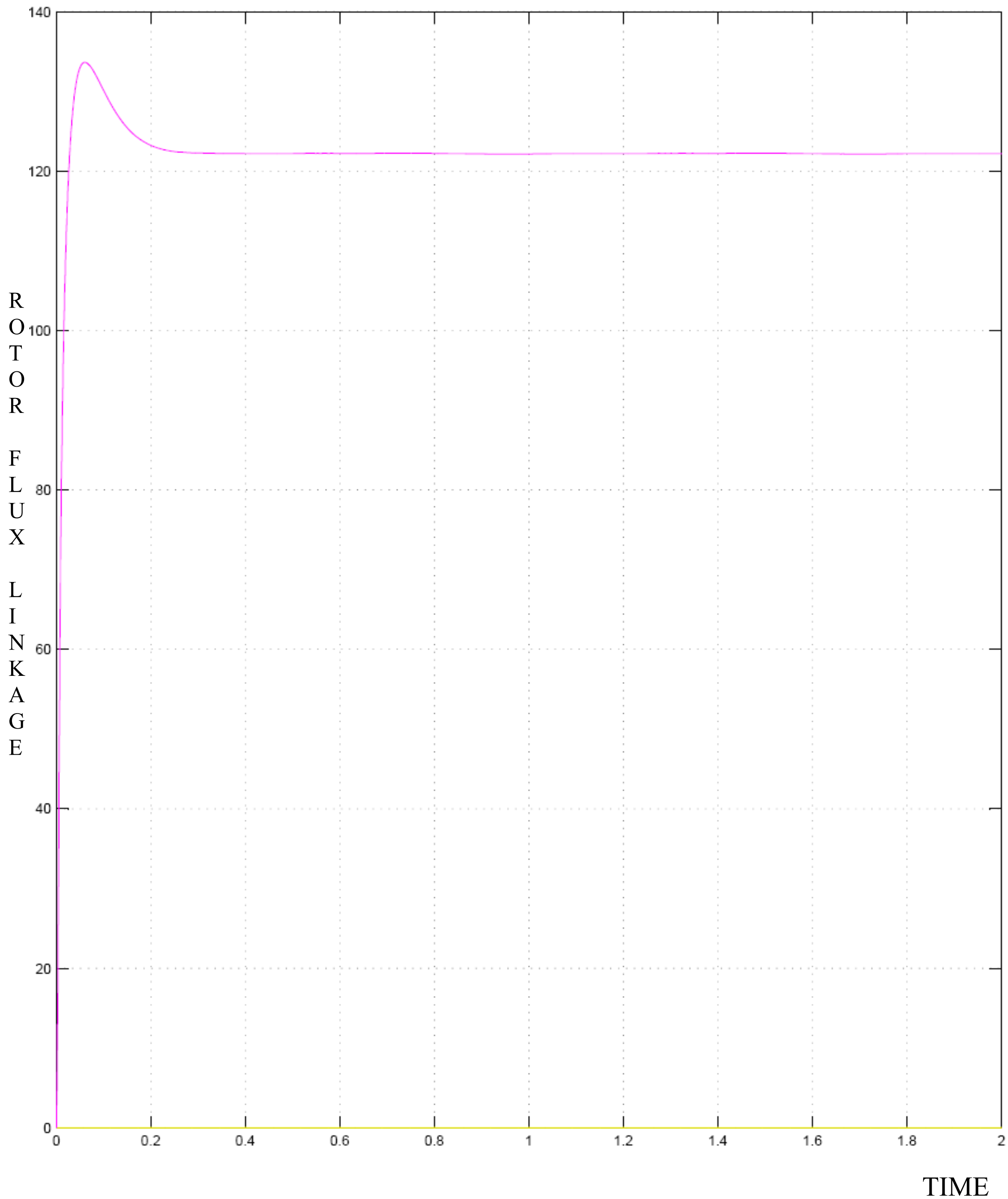


Figure 6.5 rotor flux case1

The rotor flux has been stabilized before the motor attains the full speed and the above result shows the rotor flux is independent of changes in the speed and the changes in the torque.

CASE 2:

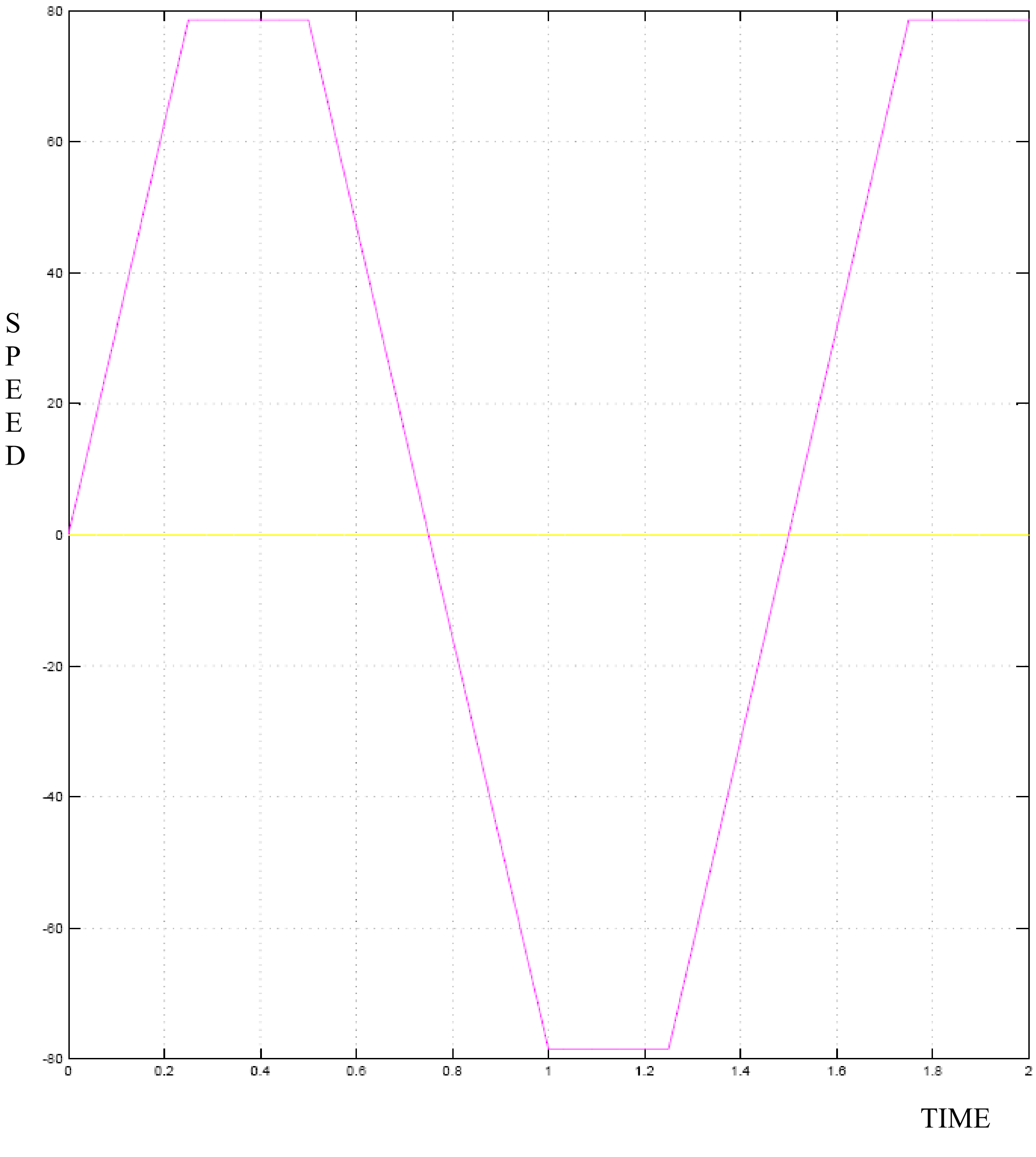


Figure 6.6 reference speed of the motor

The above figure shows the reference speed of the motor in four quadrant operation

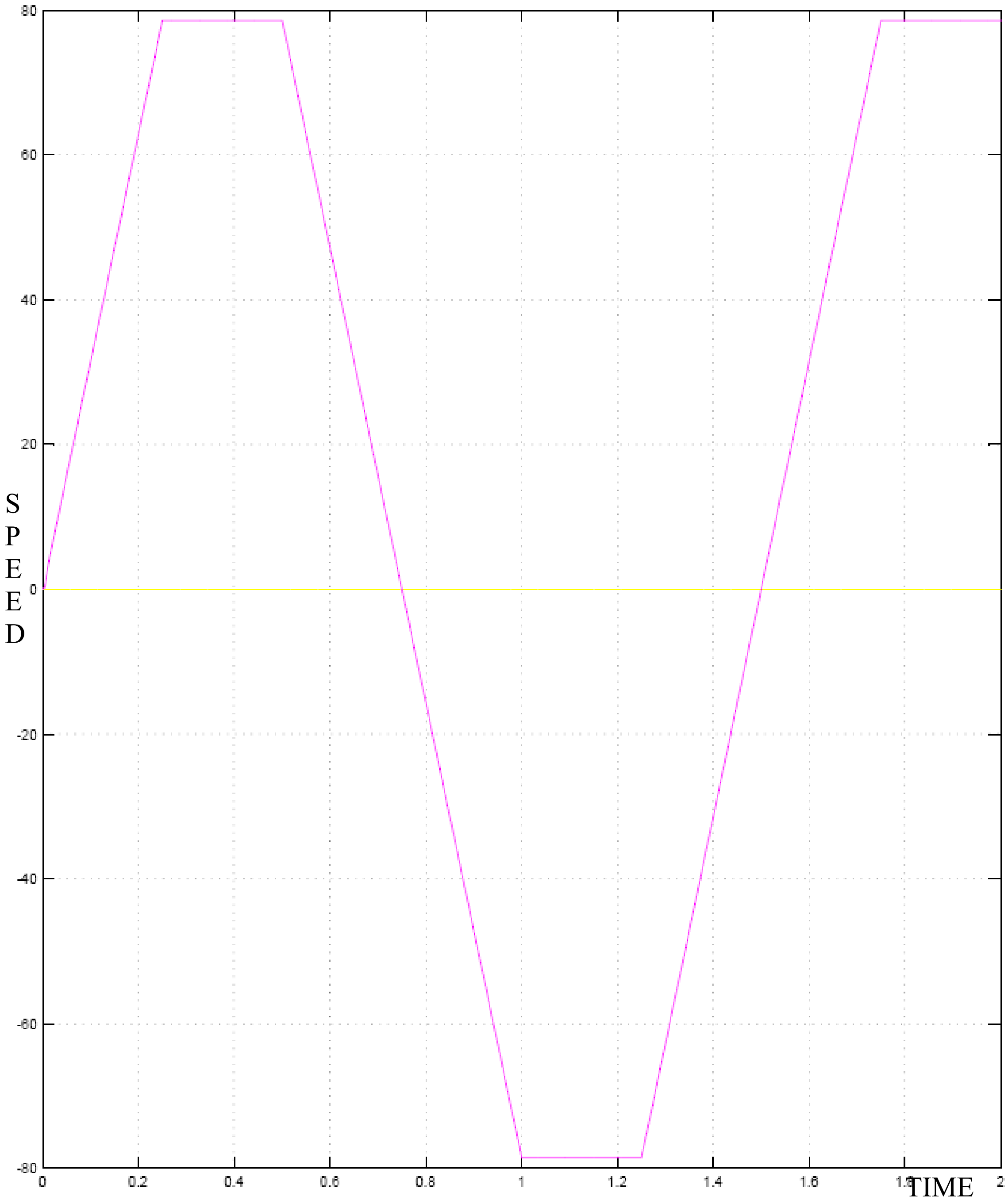


Figure 6.7 actual motor speed

The motor follows the exact reference speed in four quadrant operation also i.e. motoring and braking in both the directions

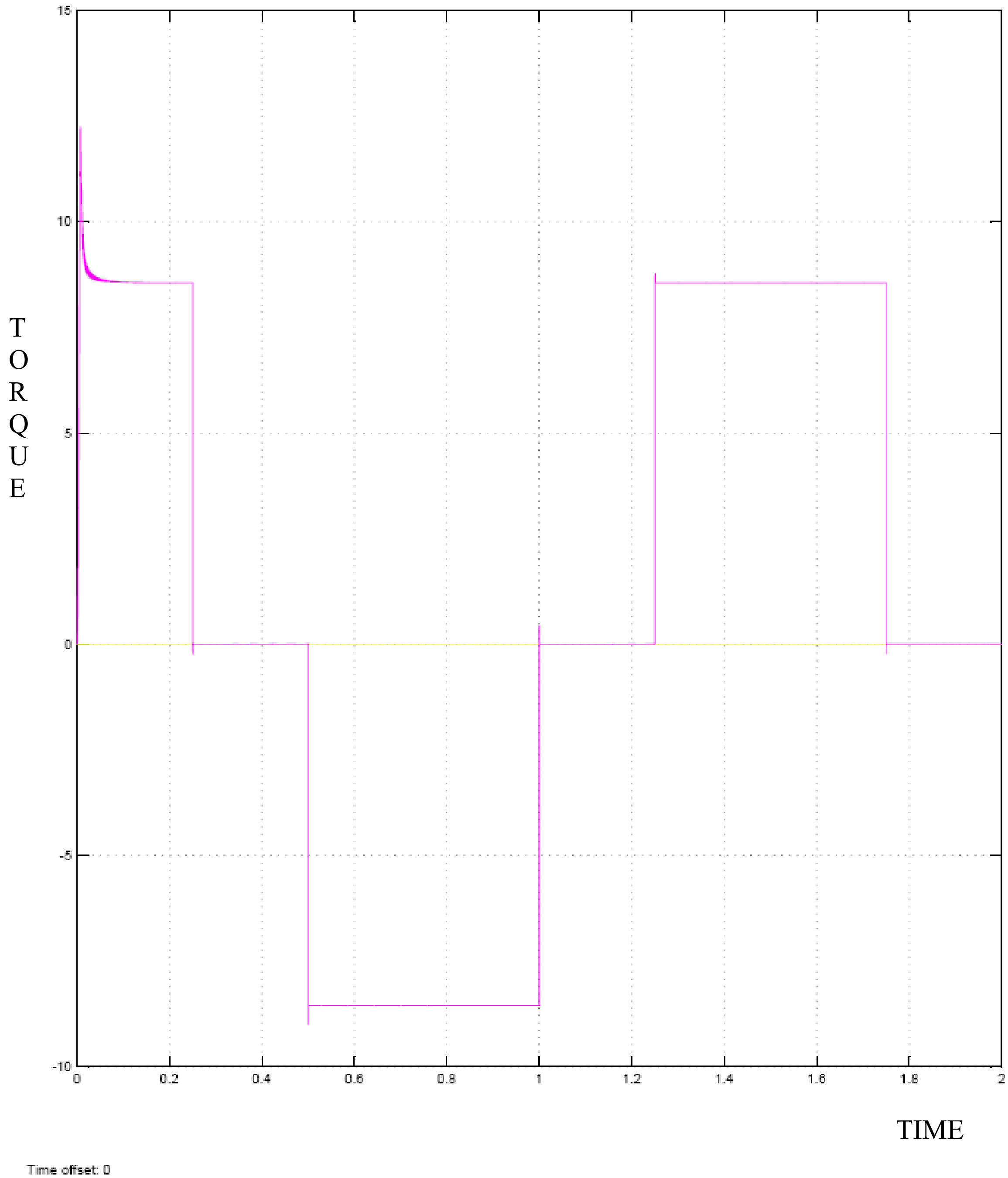


Figure 6.8 toque response

In the above response during the starting of the motor only the transient can be observed after that without any transient the motor speed can be reserved

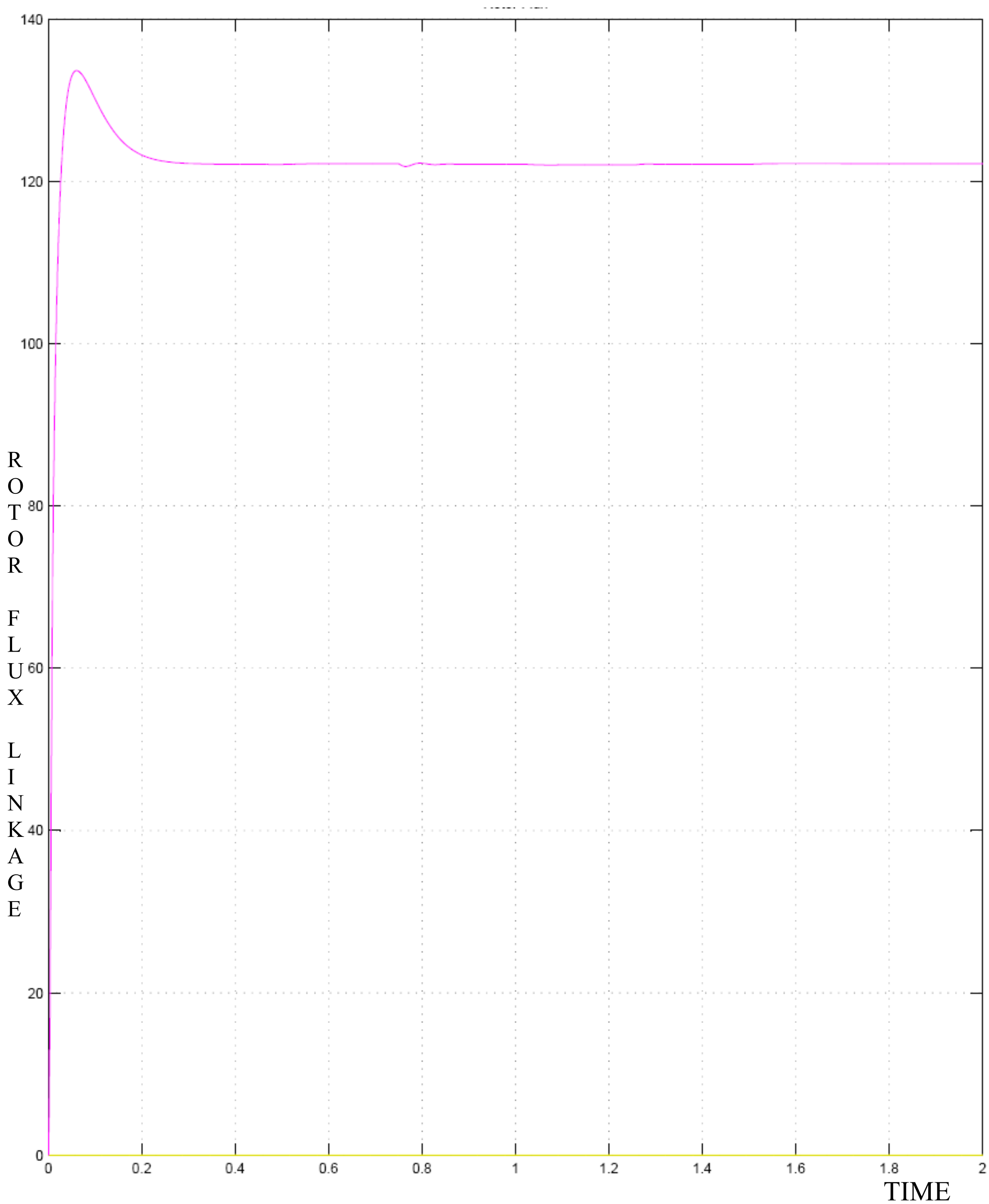


Figure 6.9 rotor flux case2

The rotor flux is constant throughout the operation i.e. four quadrant operation after once it is stabilized

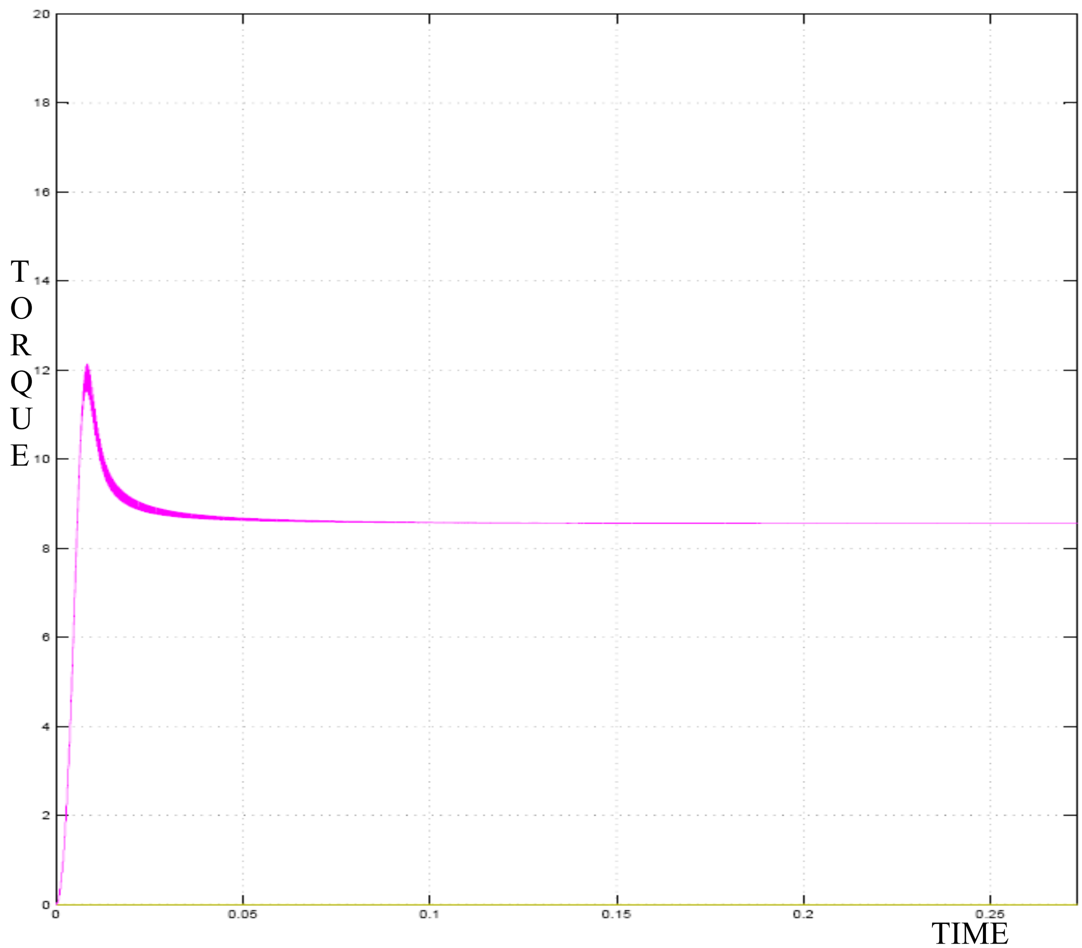


Figure 6.10 starting transient toque

The starting torque is near about 12 Nm where the full load torque is about 100Nm and this transient is stabilized in about 0.03 sec.

6.3 Conclusion:

In this chapter the motor input parameters are given working of the Simulink model is explained for different loading conditions and for four quadrant operation of the induction motor and the output results are displayed.

Chapter 7

CONCLUSION AND FUTER WORK

7.1 Conclusion:

An indirect vector controlled induction motor drive using a fuzzy logic controller has been presented. The conclusions are defined as follows.

- Torque reference computation for achieving decoupled control of the stator current components is based on the non-linear characteristics of the fuzzy controller.
- In the proposed system the transient response of the starting torque of the motor is reduced to nearly half of the previously proposed models.
- It is shown that to track command speeds and step disturbances in load torque with zero steady state error and very largely reducing settling times during starting.

7.2 Further work:

All further work is summarized schematically in the following ideas:

- The fuzzy rules used can be further tuned.

- Developments of new fuzzy controllers to achieve better performance. This new fuzzy controller should, at least take into account the following ideas:

    - Develop a completely auto adaptive controller.
    - The controller must be adaptive to any motor.
    - Try to overcome the electrical noises, which appear in any power drive.

- Study the torque ripple reduction not only with fuzzy modulators but also with multilevel converters.
- Find optimum controllers, not only for torque ripple reductions, but also for reducing EMI and for increasing energy savings form the mains.
- Quantify the real savings of reactive power obtained from the optimized stator flux reference value

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APPENDIX

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High Precision Servo Control of Induction Motor Drive using Indirect Vector Control with Fuzzy PI controller

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Abstract—**This paper presents High precise vector controlled induction motor drive using fuzzy controller in which the torque and the flux producing components are generated from a Fuzzy logic based controller. The indirect vector control method is used in which the torque and the flux are independently controlled as in the case of the dc motor. Simulated results through Sim Power Sim/Simulink (MATLAB) shows that proposed system tracks command speed and rejects the step disturbances in load torque with zero steady state error and very least transient response error which provides instantaneous torque control. The four quadrant operation is simulated through the developed controller, and the excellent results shows the superb working of fuzzy rule based controller of indirect vector control of induction motor**

I. INTRODUCTION

The application of induction motor for precise speed/torque control has been receiving growing interest due to there low cost and rugged construction. Traditional induction motor drive uses the voltage / frequency and slip frequency control [1], [9], [10], [11] are not able to achieve fast dynamic responses when subjected to impact loads and rapid changes in the reference speed. However such requirements can be met by realizing independent control of torque and machine flux, as it is commonly performed in the dc motor drives. On the other hand rotor flux is induced in an induction motor, rather than being controllable separately. Hence unlike a dc motor the rotor and the stator fluxes are not necessarily orthogonal to each other.

Vector control aims at decoupling the torque and flux producing components of the stator current under all operating speeds and load conditions [2], [12], [13], [14]. Consequently, the drive can be turned for quasi-instantaneous tracking load and reference changes. Speed information is normally applied to a PI controller. However the choice of conventional PI controller with fixed gains is strongly dependent on a linear model of the motor drive and precise speed information from tachogenerator feedback.

A fuzzy controller on the other can incorporate expert knowledge and non linear algorithms characterized by linguistic rules [3], for producing robust drive control characteristics. In [6], [7] and [16] a fuzzy logic is used to continuously adjust the gains of PI type controllers for speed tracking under varying load conditions. Although the drive performance is significantly improved mainly during the transient conditions, in [8] the two fuzzy controllers, performing coarse and fine controls, respectively, are designed to improve transient response and steady error of the induction motor drive. All the proposed controllers utilize the speed error and its derivative as input to the fuzzy inference system. This paper presents a modified two input fuzzy controller for a four quadrant induction motor drive, where one input is the speed error, and the second input is the integral of the error. Two input gains and one out gain are adjusted for meeting the speed and load tracking performance. The intrinsic nonlinear characteristics of the controller are used to compute the torque command and reference frame angle for proper field orientation. The response time and accuracy with which motor follows the speed command and corrects for load disturbances are essential parameters that determine the controller characteristics.

The paper is organized as follows: Section 2 describes the conditions necessary for vector control, section 3 presents the design of fuzzy controller, section 4 illustrates the dynamic performance of the drive system for both speed reference and load cycling modes, and section 5 concludes the simulation.

II. VECTOR CONTROL

Since the rotor flux depends on the stator current and the machine speed, the spatial orientation between the rotor and stator fields is not 90 degrees by default in an induction motor. Field oriented control aims at maintaining the orthogonal spatial angle between the rotor and the stator fields, and achieving independent control of the rotor flux. Such requirements are satisfied by controlling the magnitude and phase of the applied stator current.

Transforming the 3-phase variables of a symmetrical induction machine in a q-d axis reference frame [2] rotating synchronously with the rotor flux, the electromagnetic torque developed by the machine can be written as:

$$T_e = \frac{3PL_m}{4L'_r}\left(\lambda_{dr}{}'^{e}\, i^e{}_{qs} - \lambda_{qr}{}'^{e}\, i^e{}_{ds}\right) \tag{1}$$

Where P is the number of poles, L_m is the magnetizing inductance, $\lambda'^{e}{}_{dr}$, $\lambda'^{e}{}_{qr}$, $i^e{}_{ds}$ and $i^e{}_{qs}$ are the rotor flux and stator current direct (d) and quadrature (q) components, respectively, and

$$L'_r = L_m + L'_{lr} \tag{2}$$

L'_{lr} being the rotor equivalent leakage inductance. The rotor flux components can be expressed as:

$$\lambda'^{e}{}_{qr} = L'_r\, i'^{e}{}_{qr} + L_m i^e{}_{qs} \tag{3}$$

and

$$\lambda'^{e}{}_{dr} = L'_r\, i'^{e}{}_{dr} + L_m i^e{}_{ds} \tag{4}$$

Where $i'^{e}{}_{dr}$ and $i'^{e}{}_{qr}$ are the d-q components of the rotor current.

In a field oriented induction machine, the synchronously rotating reference frame is chosen so that its d-axis is aligned with the rotor flux. Hence $\lambda'^{e}{}_{qr}$ is zero, and from Equations (1) and (3),

$$T_e = \frac{3PL_m}{4L'_r}\lambda'^{e}{}_{dr}\, i^e{}_{qs} \tag{5}$$

And $$i'^{e}{}_{qr} = -\frac{L_m}{L'_r} i^e{}_{qs} \tag{6}$$

The corresponding positions of the stator current and rotor flux are shown in Fig 1.

The rotor induced voltage is zero for an induction machine with short circuited rotor bars. In the synchronously rotating reference frame, the q-axis and d-axis components of the rotor voltage can therefore be written as

$$v'^{e}{}_{qr} = 0 = r'_r\, i'^{e}{}_{qr} + \frac{d\lambda'^{e}{}_{qr}}{dt} + s\omega_e \lambda'^{e}{}_{dr} \tag{7}$$

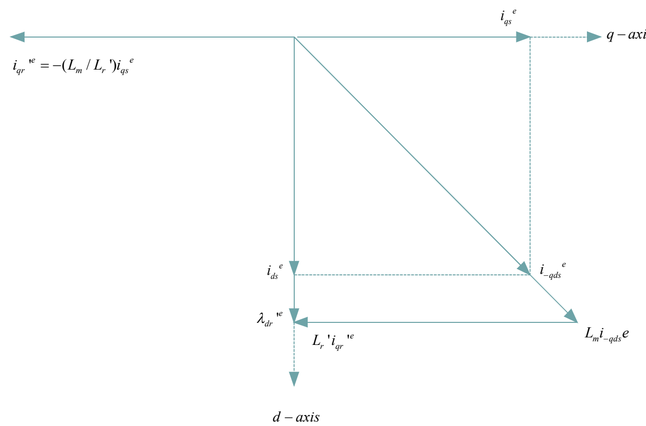

Fig 1: Motor currents and flux in synchronously rotating reference frame

$$v'^{e}{}_{dr} = 0 = r'_r\, i'^{e}{}_{dr} + \frac{d\lambda'^{e}{}_{dr}}{dt} + s\omega_e \lambda'^{e}{}_{qr} \tag{8}$$

Where $s\omega_e$ is slip frequency. The d-axis component of the rotor flux is obtained from Equations (4) and (8) as

$$\lambda'^{e}{}_{dr} = \left(\frac{L_m}{1 + \left(L'_r / r'_r\right) p}\right) i^e{}_{ds} \tag{9}$$

Where $p = \frac{d}{dt}$. With $\lambda'^{e}{}_{qr} = 0$, the steady state rotor flux is therefore

$$\lambda'^{e}{}_{dr} = L_m i^e{}_{ds} \tag{10}$$

And is aligned with the d-axis rotor current, as shown in the Fig 1. Hence, orthogonality between the rotor and stator fields can be achieved.

Equation (5) shows that the developed torque is proportional to quadrature component of the stator current and rotor flux. Equation (10) also shows that the rotor flux can be independently controlled by $i^e{}_{ds}$, as required for field orientation. Replacing $i'^{e}{}_{qr}$ by equation (6) and $\lambda'^{e}{}_{dr}$ by equation (10), the slip frequency for proper field orientation can be expressed as:

.

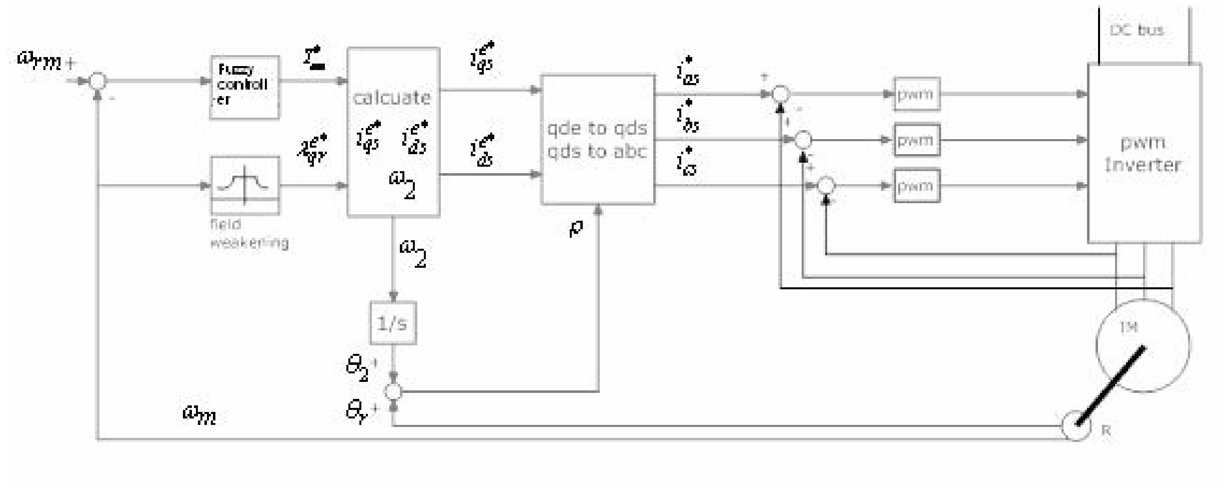

Fig. 2 Block diagram of the indirect field oriented control with fuzzy controller

$$s\omega_e = \frac{r'_r}{L'_r}\frac{i^e_{qs}}{i^e_{ds}} \tag{11}$$

From the knowledge of the rotor speed, ω_r ,the required reference frame angle, θ_{rf} for field oriented can be found as

$$\theta_{rf} = \int_0^t \left(s\omega_e + \omega_r\right)dt \tag{12}$$

A block diagram of proposed Indirect Field oriented control is shown in the Figure 2. speed information, obtained by tachogenerator feedback, enables computation of torque reference using a fuzzy controller. The later is designed to cater for non linear characteristics of the machine and estimations in the motor parameters used in the computation blocks. The drive essentially uses feed forward control ((5) to (12)) to apply the correct value of the reference frame angle so as to control the torque output and rotor flux of the machine.

The d-q current components are then transformed into 3-phase command currents (i^*_a, i^*_b and i^*_c) which are applied to the current regulated pulse width modulated (CRPWM) dc to ac inverter. The later in turn applies the correct values of stator currents to the motor for proper field orientation.

III. Fuzzy Speed Controller

The proposed fuzzy speed controller is shown in Figure 3. The error, $\Delta\omega$, between the set point and shaft speed, is used to generate the following inputs to the controller:

$$u_1 = K_P \vartriangle\omega \tag{13}$$

$$u_2 = K_I \int \vartriangle\omega dt \tag{14}$$

By proper adjustment of the member functions and the gains K_P and K_I the controller realizes a non linear PI control algorithm. The output is a torque reference, T^*_{em}, for meeting the desired speed tracking action of the drive.

The input membership functions of the fuzzy system are triangular, with increasing widths away from the origin, so as to obtain finer control action near the set point. In the fuzzification process, these membership functions crisp inputs, $u_1$and u_2, into fuzzy subsets with the following linguistic variables: Positive Big (PB), Positive Medium(PM), Positive Small (PS), Zero (ZE), Negative Small (NS), Negative Medium (NM) and Negative Big (NB). The knowledge base consists of 49 rules for determining the result of the fuzzy implication process, as shown in the Table 1. for instance, the first entry in the rule base reads

IF u_1 is “NB” AND u2 is “NB” then u3 is “NB”.

Since a finite crisp output is required, the output membership function are designed to follow a uniform distribution without end point saturation. In this paper, the *Min* operator is used for computing the premise and implication results, while the *Max* operator is used for aggregation process. The resulting overall fuzzy set is transformed into a real torque command value using Center of Gravity (CoG) defuzzification, which provides adequate compromise between accuracy and computational effort [5]. The input output mapping of the fuzzy controller is shown in the figure 4. the controller surface is almost linear for small disturbances about zero error.

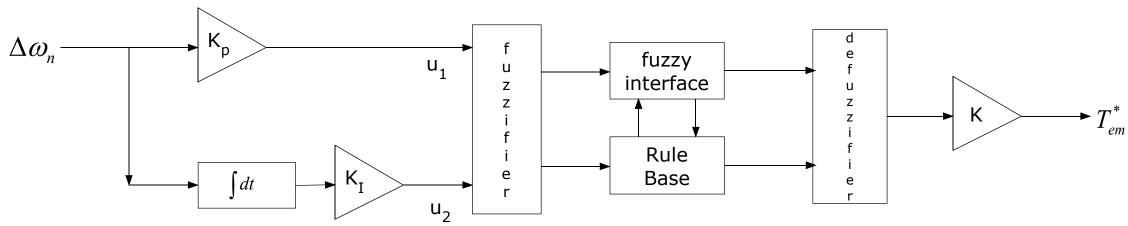

Fig 3 Fuzzy controller block diagram

Table 1: Fuzzy Rule Base

u1\u2	**NB**	**NM**	**NS**	**ZE**	**PS**	**PM**	**PB**
NB	NB	NB	NB	NB	NM	NS	ZE
NM	NB	NB	NB	NM	NS	ZE	PS
NS	NB	NB	NM	NS	ZE	PS	PM
ZE	NB	NM	NS	ZE	PS	PM	PB
PS	NM	NS	ZE	PS	PM	PB	PB
PM	NS	ZE	PS	PM	PB	PB	PB
PB	ZE	PS	PM	PB	PB	PB	PB

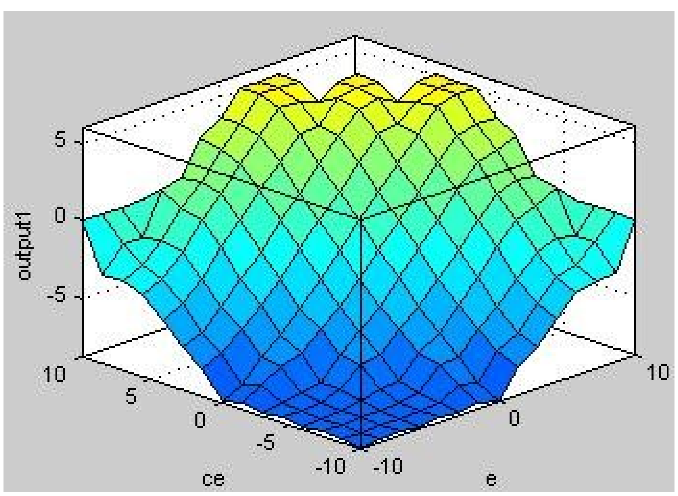

Figure 4 Surface Plot

IV. Result and Discussion

The gains K_P and K_I of the fuzzy controller were adjusted iteratively until the desired speed tracking performance were obtained.

The parameters used in the simulation of the proposed drive as follows;

Rated line to line voltage: 220V, 50Hz
Power rating: 15kVA (4pole)
Stator referred Resistances: r_s =0.10 Ω; r_s ' = 0.08 Ω
Stator referred leakage inductance: $L_s = L_r' = 725\mu H$
Magnetizing inductance L_m =18.6mH
Rotor inertia, J=2.6 kgm^2

Case 1: To investigate the load torque rejection capability of the drive, the speed is first ramped from 0 to 157 rad/s, as shown in the Figure 5. At t=0.75sec., a step load torque corresponding to the rated value is applied to the rotor shaft. The load on the rotor shaft is next reduced to 50% at the time t = 1sec. in both the cases, the motor torque response is instantaneous and the speed is restored back to the reference value in minimal time. the transient response during the start up is shown in the figure 6. this response is very much reduced.

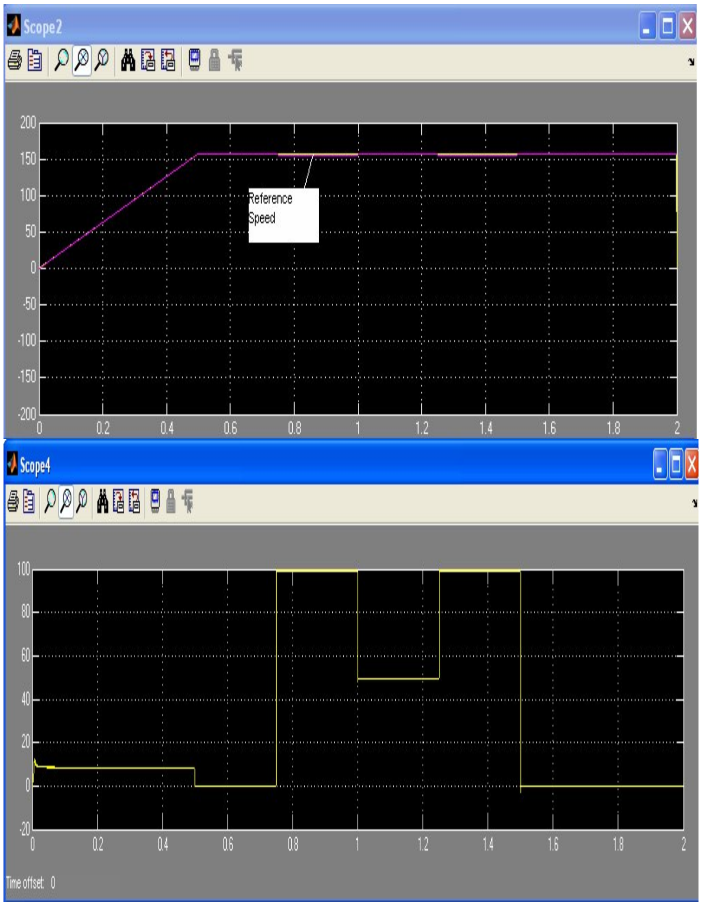

Fig. 5 The above one shows the speed response and the below one shows the torque response under loading conditions

Transient Response:

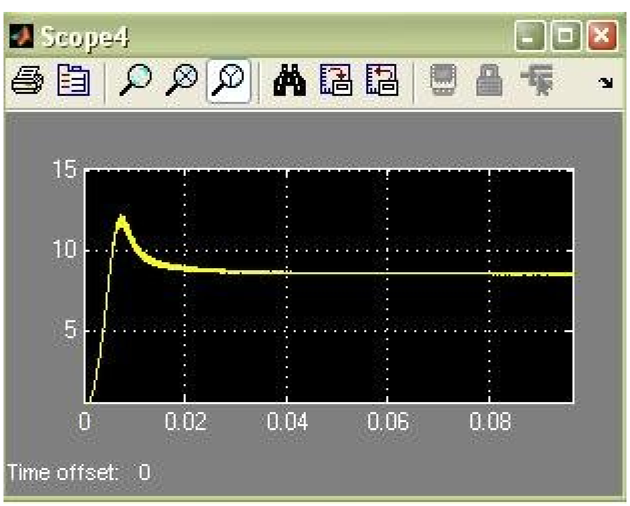

Fig 6: Transient torque during starting

Case 2: In the second case Figure 7. the reference speed is ramped up to 79rpm in the forward motoring mode, followed by linear deceleration to zero, a ramp in the opposite direction and finally reverse braking to standstill. In all four modes, the motor speed is seen to follow the reference with negligible error.

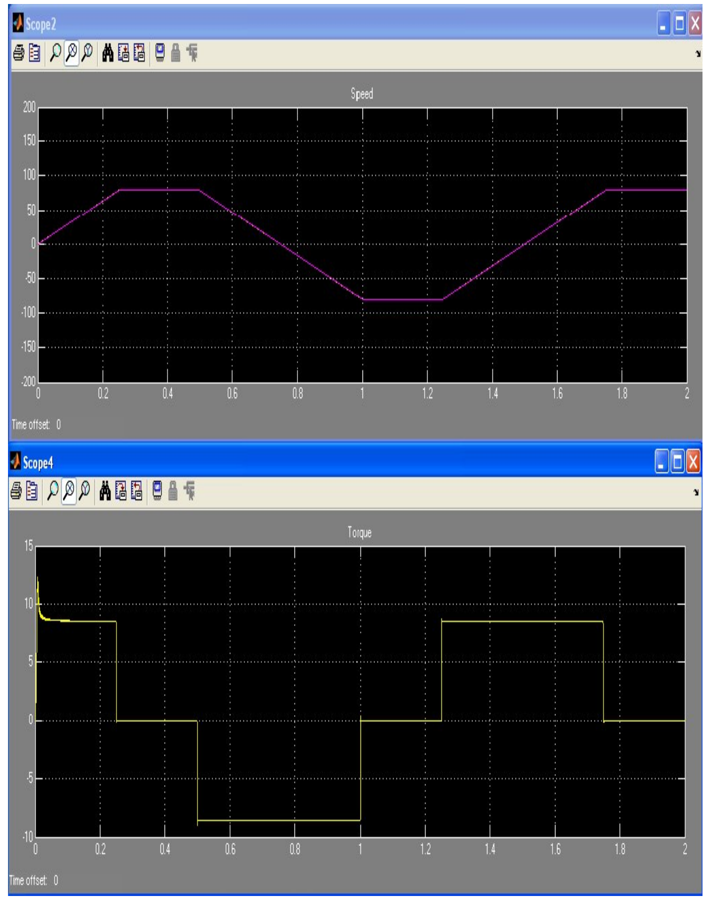

Fig. 7 The above one shows the speed response and the below one shows the torque response under no load

V. Conclusion

An indirect vector controlled induction motor drive using a fuzzy logic controller has been presented. Torque reference computation for achieving decoupled control of the stator current components is based on the non-linear characteristics of the of the fuzzy controller. In the proposed system the transient response of the starting torque of the motor is reduced to nearly half of the previously proposed models and it is shown that to track command speeds and step disturbances in load torque with zero steady state error and very largely reducing settling times during starting.

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