INTELLIGENT CONTROL OF INDUCTION MOTOR

A Project Report

Submitted by

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BONAFIDE CERTIFICATE

Certified that this project report entitled "**Intelligent Control of Induction Motor**" is the bonafide work of

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Who carried out the project work under my supervision.

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Coimbatore has participated in the SECOND NATIONAL CONFERENCE on "CUTTING EDGE TECHNOLOGIES IN POWER CONVERSION AND INDUSTRIAL DRIVES", PCID-2006 held on 24-25, March 2006 and presented a paper titled "Intelligent Control of Induction Motor"

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ABSTRACT

AC Motors, Particularly the squirrel-cage Induction motor (SCIM), enjoy several inherent advantages like simplicity, reliability, low cost and virtually maintenance-free electrical drives. However, for high dynamic performance industrial applications, their control remains a challenging problem because they exhibit significant non-linearities and many of the parameters, mainly the rotor resistance, vary with the operating conditions. Field orientation control (FOC) or Vector control of an Induction motor achieves decoupled torque and flux dynamics leading to independent control of the torque and flux as for a separately excited DC motor. FOC methods are attractive but suffer from one major disadvantage: they are sensitive to motor parameter variations such as the rotor time constant and an incorrect flux measurement or estimation at low speeds. Consequently, performance deteriorates and a conventional controller such as a PID is unable to maintain satisfactory performance under these conditions.

In this project two control methods are introduced with Indirect Field-Oriented Controller for Controlling Induction Motor and their output is compared with the conventional PI controller. In the first approach fuzzy logic controller (FLC) is developed. This follows the interpretation of linguistic IF-THEN rules. The parameters of the Fuzzy Logic Controller does not affected by the machine parameters because the fuzzy logic do not need any machine parameters. So the Robustness of the system is improved.

The Second design approach is based on the well known Model Reference Adaptive Control (MRAC). The plant's response is forced to track the response of the reference model and by controlling the output error the speed of the motor is controlled.

These controllers are evaluated under simulations for a variety of operating conditions of the drive system and the results demonstrate the ability of the control structures to improve the performance of the drive system.

ஆய்வுச் சுருக்கம்

மாறுதிசை மின் இயந்திரங்கள் அவற்றின் நம்பகத்தன்மை, விலை குறைவு மற்றும் குறைந்த பராமரிப்பு போன்ற காரணங்களால் தொழிற்சாலைகளில் மிகவும் பயன்படுத்தப்படும் இயந்திரங்களாக உள்ளன. ஆனால் அதிக இயக்க வேறுபாடுகளைக் கொண்ட தொழிற்சாலைகளில் இவற்றின் செயல்பாடுகளைக் கட்டுப்படுத்துதல் ஒரு கடினமான மற்றும் சவாலான காரியமாகும். ஏனெனில், இந்த வகையான இயந்திரங்கள் அதிகமாக எதிர்பாராத வெளியீடுகளால் பாதிக்கப்படுகின்றன. முக்கிமாக ரோட்டார் மின்தடை இயக்க முறைகளுக்கு ஏற்ப மாற்றமடைகிறது.

காந்தப் புலம் சார்ந்த கட்டுப்பாட்டு முறைகள் (குடீஊ) தனித்தனியான டார்க் மற்றும் காந்தப் புலம் கட்டுப்பாட்டு முறைகள் மூலம் ஒன்றையொன்று சார்ந்திராத கட்டுப்பாட்டு முறைகளை உறுவாக்குகின்றன. இந்த வகையான செயல்பாட்டினால் மாறுதிசை மின் இயந்திரங்கள், நேர்மின் இயந்திரங்களைப் போல செயல்படுகின்றன.

ஆனால் காந்தப் புலம் சார்ந்த கட்டுப்பாட்டு முறைகள் (குடீஊ)இ இயந்திரச் செயல்பாட்டில் ஏற்படும் மாற்றங்களினால் அதிகமாக பாதிக்கப்படுகின்றன. எனவே பழமையான கட்டுப்பாட்டு முறைகளின் செயல்பாடுகளும் பாதிக்கப்படுகின்றன.

இந்த ஆய்வில் இரண்டு புது வகையான கட்டுப்பாட்டு முறைகளின் செயல்பாடுகள் ஆராயப்படுகின்றன. ஒன்று இயந்திரங்களின் வேகத்தை ' பஸி தத்துவம் மூலம் கட்டுபடுத்துதல். இவற்றின் மூலம் இயந்திரங்கள் நிலைப்புத்தன்மை அதிகரிக்கும்.

இரண்டு எம். ஆர். ஏ. சி. முறையை அடிப்படையாகக் கொண்ட முறையாகும். ஒரு ஐடியல் மாடல் இயந்திரத்துடன் இணைக்கப்பட்டு அதன் வெளியீட்டுப் பிழைகளை பூஜ்யமாக்குவதன் மூலம் வேகத்தைக் கட்டுப்படுத்துவதாகும்.

இரண்டு வகையான கட்டுப்பாட்டு முறைகளின் மாதிரிகளும் மேட்லேப் என்ற மென்பொருள் மூலம் சிமுலேஷன் செய்யப்பட்டு, அவற்றின் செயல்பாடுகள் ஆராயப்பட்டு வெளியீடுகள் ஒப்பிடப்பட்டுள்ளன.

ACKNOWLEDGEMENT

I humbly submit all the glory and thanks to the almighty for showering the blessings and giving the necessary wisdom for accomplishing this project.

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I would like to extend a special thanks to my friends, teaching and non teaching staffs and my parents who have directly and indirectly contributed to the success of this project.

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LIST OF SYMBOLS AND ABBREVIATIONS

P_n	Nominal Power
T_e, T_L	Electromagnetic and Load Torques
J	Rotor Inertia
p	Pairs of Poles
ω	Motor Speed
R_s, R_r	Stator and Rotor Resistances
L_s, L_r	Stator and Rotor Inductance
L_m	Mutual Inductance
T_r	Rotor time constant
$[i_{ds}^s \; i_{qs}^s]$	d- and q- axis stator currents
$[i_{dr}^s \; i_{qr}^s]$	d- and q- axis rotor currents
$[V_{ds}^s \; V_{qs}^s]$	d- and q- axis stator voltages
$[i_{ds}^{e*} \; i_{qs}^{e*}]$	Stator current references
$[\Phi_{dr}^s \; \Phi_{qr}^s]$	Flux linkages in the stator reference frame
SCIM	Squirrel cage Induction Motor
FOC	Field oriented control
PI	Proportional- Integral
FLC	Fuzzy logic controller
NB	Negative Big
NM	Negative Medium
NS	Negative Small
Z	Zero
PS	Positive Small
PM	Positive Medium
PB	Positive Big
MRAC	Model Reference Adaptive Control
F	Viscous friction constant
J	Moment of inertia
T_s	Sampling time
K_T	Torque constant

1. INTRODUCTION

1.1. NEED FOR THE PROJECT

The squirrel-cage Induction motors (SCIM) are used in most of the industries due to their simplicity, reliability, low cost and virtually maintenance-free advantages. However, for high dynamic performance industrial applications, their control remains a challenging problem because they exhibit significant non-linearities and many of the parameters, mainly the rotor resistance, vary with the operating conditions. Field orientation control (FOC) or Vector control of an Induction motor achieves decoupled torque and flux dynamics leading to independent control of the torque and flux as for a separately excited DC motor. FOC methods are attractive but suffer from one major disadvantage: they are sensitive to motor parameter variations such as the rotor time constant and an incorrect flux measurement or estimation at low speeds. Consequently, performance deteriorates and a conventional controller such as a PID is unable to maintain satisfactory performance under these conditions. So we go for the new type of controllers to improve the dynamic response and robustness of the motor.

1.2. OBJECTIVE OF THE PROJECT

The aim of my project is to control the speed and to improve the dynamic performance and robustness of induction motor. Already existing scalar control has many disadvantages. The inherent coupling effect in scalar control gives sluggish dynamic response. In order to overcome this effect vector control with fuzzy logic controller and Model Reference Adaptive Controller is used in the project.

1.3. ORGANISATION OF THE REPORT

CHAPTER-1:

This chapter contains the information about the main objective and need for the project.

CHAPTER-2:

This chapter contains information about the Constructional Details, Scalar control Methods, Applications, Advantages and Disadvantages of the Squirrel cage Induction Motor.

CHAPTER-3:

This chapter contains information about the Basic principle of the Vector control Method and the dq- model of Induction Motor with the basic transformations used in vector control along with the direct vector method.

CHAPTER-4:

This chapter contains information about the Indirect Vector Control Method along with Hysteresis Band current Control PWM Method.

CHAPTER-5:

This chapter deals about the function of Fuzzy Logic Controller (FLC) with an Introduction about the Fuzzy Logic.

CHAPTER-6:

This chapter deals about the Function of Model Reference Adaptive Controller (MRAC) with the basic Introduction of Adaptive control Techniques.

2. INDUCTION MOTOR

2.1 BASIC PRINCIPLE SCIM

The AC induction motor is a rotating electric machine designed to operate from a three-phase source of alternating voltage. The stator is a classic three phase stator with the winding displaced by 120°. The most common type of induction motor has a squirrel cage rotor in which aluminium conductors or bars are shorted together at both ends of the rotor by cast aluminium end rings. When three phase currents flow through the three symmetrically placed windings, a sinusoidally distributed air gap flux generating the rotor current is produced. The interaction of the sinusoidally distributed air gap flux and induced rotor currents produces a torque on the rotor. The mechanical angular velocity of the rotor is lower than the angular velocity of the flux wave by so called slip velocity. In adjustable speed applications, AC motors are powered by inverters.

2.2 CONSTRUCTION

2.2.1 Stator-Design

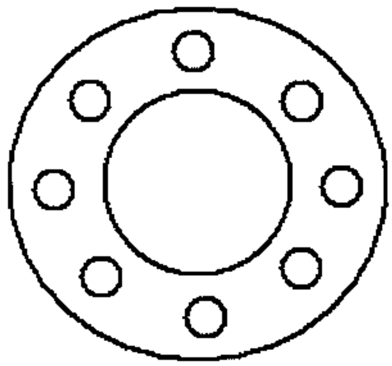
The stator is the outer body of the motor which houses the driven windings on an iron core. In a single speed three phase motor design, the standard stator has three windings, while a single phase motor typically has two windings. The stator core is made up of a stack of round pre-punched laminations pressed into a frame which may be made of aluminium or cast iron. The laminations are basically round with a round hole inside through which the rotor is positioned. The inner surface of the stator is made up of a number of deep slots or grooves right around the stator. It is into these slòts that the windings are positioned. The arrangement of the windings or coils within the stator determines the number of poles of the squirrel cage induction motor.

A standard bar magnet has two poles, generally known as North and South. Likewise, an electromagnet also has a North and a South pole. As the induction motor Stator is essentially like one or more electromagnets depending on the stator windings, it also has poles in multiples of two. i.e. 2 pole, 4 pole, 6 pole, 8 pole etc.

The winding configuration; slot configuration and lamination steel all have an effect on the performance of the motor. The voltage rating of the motor is determined by the number of turns on the stator and the power rating of the motor is determined by the losses which comprise copper loss and iron loss, and the ability of the motor to dissipate the heat generated by these losses. The stator design determines the rated speed of the motor and most of the full load, full speed characteristics.

2.2.2 Rotor Design

The rotor core is cylindrical and slotted on its periphery. The rotor consists of uninsulated copper or aluminium bars called rotor conductors. The bars are placed in the slots. These bars are permanently shorted at each end with the help of conducting copper ring called end ring. The bars are usually brazed at each end to provide good mechanical strength. The entire structure looks like a cage, forming a closed electrical circuit. So the rotor is called squirrel cage rotor. The construction is shown in the figure 1.2.



Figl.1 Symbolic representation of squirrel cage rotor

As the bars are permanently shorted to each other through end ring, the entire rotor resistance is very small. Hence this rotor is also called short circuited rotor. As rotor itself is short circuited, no external resistance can have any effect on the rotor resistance. Hence no external resistance can be introduced in the rotor circuit. So slip ring and brush assembly is not required. Hence the construction of this type of rotor is simple.

Fan blades are generally provided at the ends of the rotor core. This circulates the air through the machine while operation, providing the necessary cooling. The air gap between stator and rotor is kept uniform as small as possible.

2.3 EQUIVALENT CIRCUIT

The induction motor can be treated essentially as a transformer for analysis. The induction motor has stator leakage reactance, stator copper loss elements as series components, and iron loss and magnetising inductance as shunt elements. The rotor circuit likewise has rotor leakage reactance, rotor copper (aluminium) loss and shaft power as series elements. The transformer in the centre of the equivalent circuit can be eliminated by adjusting the values of the rotor components in accordance with the effective turns ratio of the transformer. From the equivalent circuit and a basic knowledge of the operation of the induction motor, it can be seen that the magnetising current component and the iron loss of the motor are voltage dependent, and not load dependent.

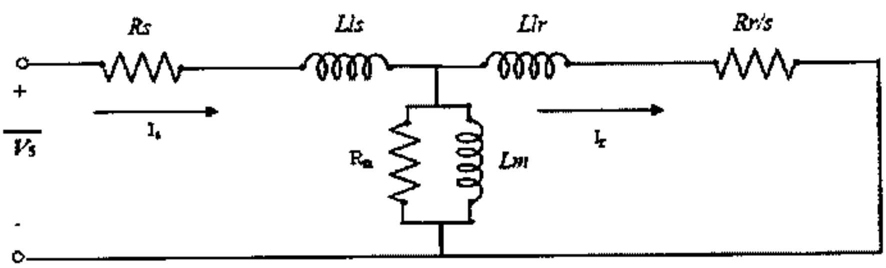


Fig.1.2 Per phase Equivalent circuit

Additionally, the full voltage starting current of a particular motor is voltage and speed dependant, but not load dependant. The magnetising current varies depending on the design of the motor. For small motors, the magnetising current may be as high as 60%, but for large two pole motors, the magnetising current is more typically 20 - 25%. At the design voltage, the iron is typically near saturation, so the iron loss and magnetising current do not vary linearly with voltage with small increases in voltage resulting in a high increase in magnetising current and iron loss.

2.4 SALIENT FEATURES

- Medium construction complexity, multiple fields on stator, cage on rotor
- High reliability (no brush wear), even at very high achievable speeds
- Medium efficiency at low speed, high efficiency at high speed

- Driven by multi-phase Inverter controllers
- Sensorless speed control possible
- Low cost per horsepower
- Higher start torque than for single-phase, easy to reverse motor

2.5 APPLICATION

- Compressors
- Air conditioning units
- Pumps
- simple industrial drives
- Electric cars
- Industrial machines

2.6 INDUCTION MOTOR CONTROL

There are two types of controls used in induction motors. They are

1. Scalar control
2. Vector control

2.6.1 SCALAR CONTROL

Scalar control is due to magnitude variation of control variables only and disregards the coupling effect in the machine. The scalar control techniques are

1. Voltage fed inverter control
2. Current fed inverter control

VOLTAGE FED INVERTER CONTROL

The voltage fed inverter controls are

1. open loop volts/Hz control
2. Energy conservation effect by variable frequency drive
3. Speed control with slip regulation
4. Speed control with torque and flux control
5. Current control voltage fed inverter drive

CURRENT FED INVERTER CONTROL

The Current fed inverter controls are

1. Independent current and frequency control
2. Speed and flux control in current fed inverter drive
3. Volts/Hz control of current fed inverter drive.

   Among all above voltage fed invert controls are the most popular technique used for controlling induction motor is open loop V/F control

2.6.2. OPEN LOOP Volts/Hz CONTROL

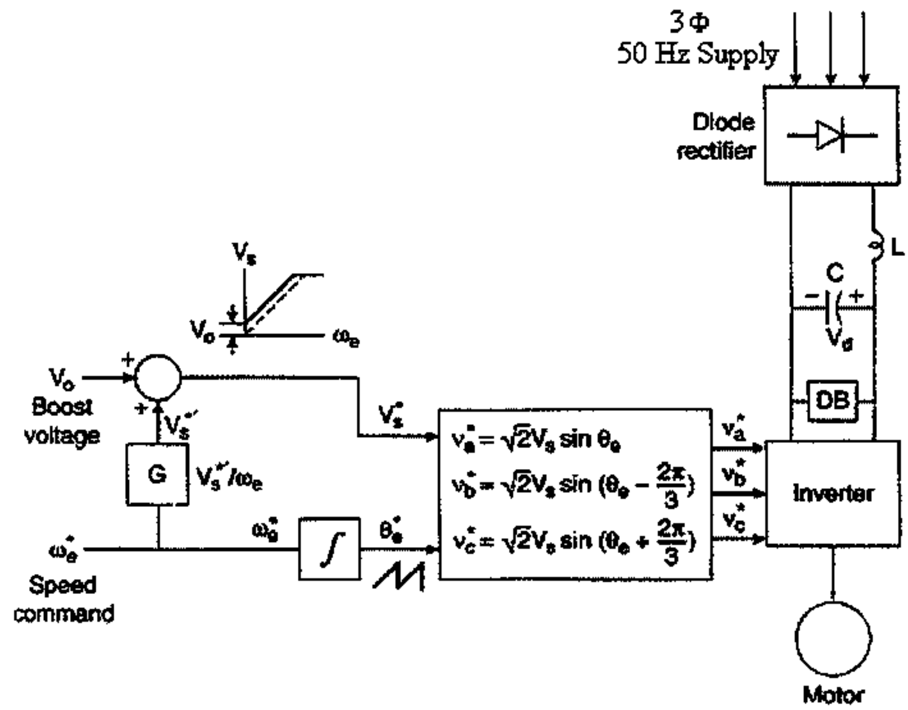


Fig.1.3 Open loop V/F speed control

Fig 1.5 shows the block diagram of open loop V/F speed control method. The power circuit consist of diode rectifier with a single phase or three phase supply, LC filter and PWM voltage fed inverter. Ideally, no feed back signals are needed for the control. The frequency ω_e is the primary control variable because it is approximately equal to speed ω_r neglecting the small slip frequency ω_{sl} of the machine. The phase voltage V_s^* command is directly generated from the frequency command by the gain factor G as shown, so that the flux Ψ_s remains constant. If the stator resistance and leakage of the inductance of the machine are neglected, the flux will also correspond to the air gap flux Ψ_m or rotor flux Ψ_r. As the frequency becomes small at low speed, the stator resistance tends to absorb the major amount of the stator voltage thus weakening the flux. The boost voltage V_o is added so the rated flux and corresponding full load torque become available down to zero speed. Note that the effect of boost voltage becomes negligible at higher frequencies. The ω_e^* signal is integrated to generate the angle signal θ_e, and the corresponding sinusoidal phase voltages (v_a^*, v_b^*, v_c^*) are generated by the expressions shown in fig1.3. The PWM controller is merged with the inverter block

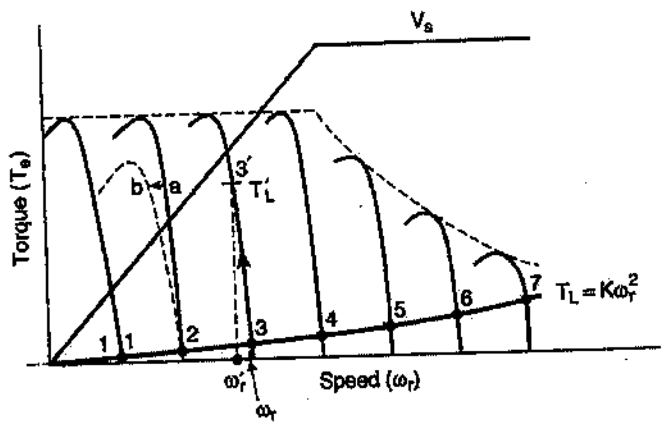


Fig.1.4 Torque and speed characteristics

Fig 1.5 shows the drives steady state performance on a torque speed plain with a fan load (or) pump type load $(T_L = k\omega_r^2)$. As frequency is gradually increased, the speed also increases almost proportionally, as indicated by points 1,2,3,4 etc. The operation can be continued smoothly in the feel weakening region in the supply voltage saturates.

Now consider the load torque and line voltage variation effects. If the initial operating point is 3 and the load torque increased to T_L' for the same frequency command, the speed will droop from ω_r to ω_r' This droop the small, particularly the high efficiency machine, and is early tolerated for a pump are fan type drive ware precision speed control is not necessary. Assume now that the operating point in another torque speed curve. If the AC line voltage decreases that lowers the machine terminal voltage.

The speed will droop correction in an open loop control can be achieved by adding an estimated slip signal with the frequency command. Performance is similar to that power factor acceleration performance

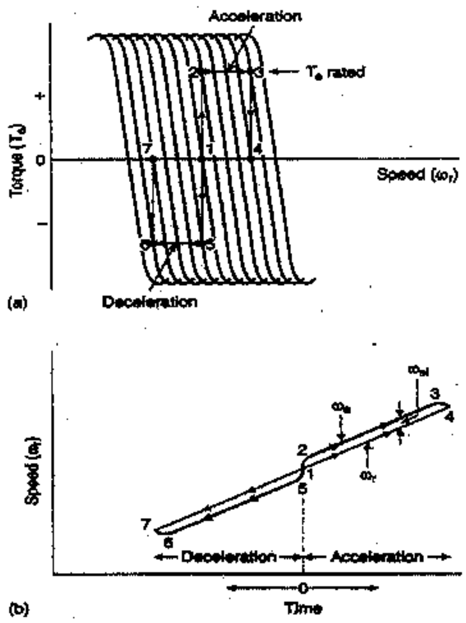


Fig 1.5 Speed Torque and time characteristics of Induction motor

2.7. ADVANTAGES & DISADVANTAGES OF SCALAR CONTROL

Advantages

1. It is simple and straight forward
2. It is very easy to implement

Disadvantages

1. The large abrupt change in frequency leads to instability of the machine.
2. The desired output speed control cannot be maintained precisely because of open loop control.

3. VECTOR CONTROL OF INDUCTION MOTORS

3.1 INTRODUCTION

Scalar control technique of voltage fed and current fed inverter drives are simple to implement, but the inherent coupling effect gives sluggish response and the system is easily prone to instability because of high order system effect. These problems can be solved by vector control. In this a squirrel cage induction motor can be controlled like a separately excited dc motor, with high performance.

Because of dc machine like performance, vector control is also known as decoupling, orthogonal, Trans vector control or Field Oriented Control (FOC).Vector control is applicable to both induction and synchronous motor drives. There are two types of vector control namely direct vector control and indirect vector control. I am doing project in direct vector control.

3.2 BASIC PRINCIPLE OF FIELD-ORIENTED CONTROL

With a vector control, an induction motor can operate as a separately excited dc motor. In a dc machine, the developed torque is given by

$$\mathbf{T_d = K_t I_a I_f}$$

Where I_a is the armature current and I_f is the field current. The construction of a dc machine is such that the field linkage ψ_a produced by I_a. These flux vectors that are stationary in space are orthogonal or decoupled in nature. As a result, a dc motor has fast transient response. However, an induction motor cannot give such fast response due to its inherent coupling problem. However, an induction motor can exhibit the dc machine characteristic if the machine is controlled in a synchronously rotating frame ($d^e - q^e$), where the sinusoidal machine variables appear as dc quantities in the steady state.

Fig 3.1 shows an inverter-fed induction motor with two control current inputs: i^*_{ds} and i^*_{qs} are the direct-axis component and the quadrature-axis component of the stator current, respectively, in a synchronously rotating reference frame. With vector control, i^*_{ds} is analogous to the field current I_f and i^*_{qs} is analogous to armature current I_a of a dc motor. Therefore, the developed torque of an induction motor is given by

$$\mathbf{T_d = K_m \psi^* I_f = K_t I_{ds} I_{qs}}$$

Where ψ^* is the absolute peak value of the sinusoidal space flux linkage vector ψ_r.

I_{ds} is the field component.

I_{qs} is the armature component.

Fig 3.1 shows the space vector diagram for vector control: i^*_{ds} is oriented (or aligned) in the direction of rotor flux λ_r, and i^*_{qs} must be perpendicular to it under all operating conditions. The space vectors rotate synchronously at frequency ω_s. Thus, the vector control must ensure the correct orientation of the space vectors and generate the control input signals.

The implementation of the vector control is shown in fig 3.2. The inverter generates currents i_a, i_b and i_c in response to the corresponding command currents i_a^*, i_b^* and i_c^* are converted to i^s_{ds} and i^s_{qs} components by three-phase to two-phase transformation. These are then converted to synchronously rotating frame (into I_{qs} and i_{qs} components) by the unit vector components $\cos\theta_s$ and $\sin\theta_s$ before applying them to the machine. The machine is represented by internal conversions into the $d^e - q^e$ model.

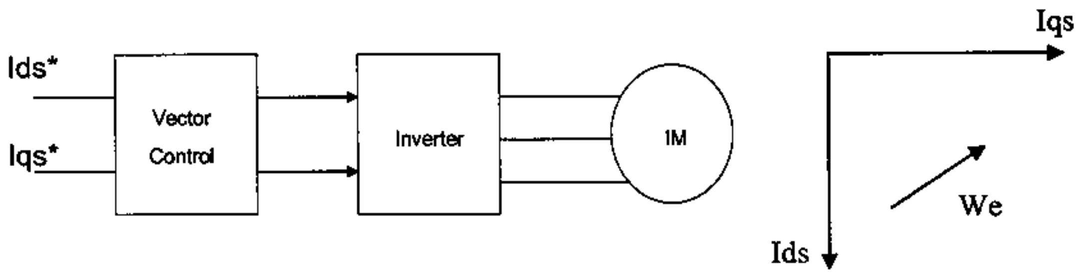


Fig 3.1. Block Diagram

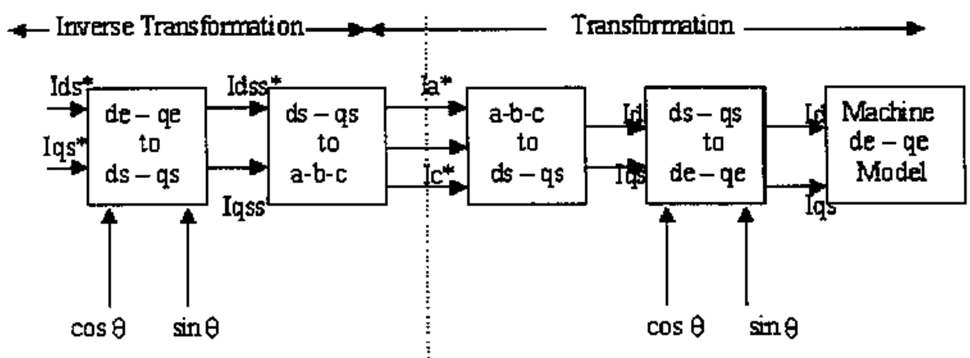


Fig 3.2 Vector Control Implementation

The controller makes two stages of inverse transformation so that the line control currents i^*_{ds} and i^*_{qs} correspond to the machine currents I_{ds} and I_{qs} respectively. In addition, the unit vector ($\cos\theta_e$ and $\sin\theta_e$) ensures correct alignment of i_{ds} current with flux vector ψ_r and i_{qs} current is perpendicular to it. It is important to note that the transformation and inverse transformation ideally do not incorporate any dynamics. Therefore, the response to I_{ds} and I_{qs} is instantaneous except for any delays due to computational and sampling times.

3.3 dq MODEL OF INDUCTION MOTOR

3.3.1 Axes Transformation

Consider the symmetrical three phase induction machine with stationary as-bs-cs axes at $2\pi/3$ angle apart as in figure.3.3. Our aim to transform the three-phase stationary reference frame (a_s-b_s-c_s) variables into two-phase stationary reference frame (d^e- q^e) variables and then transform these to synchronously rotating reference frame (d^e - q^e), and vice-versa.

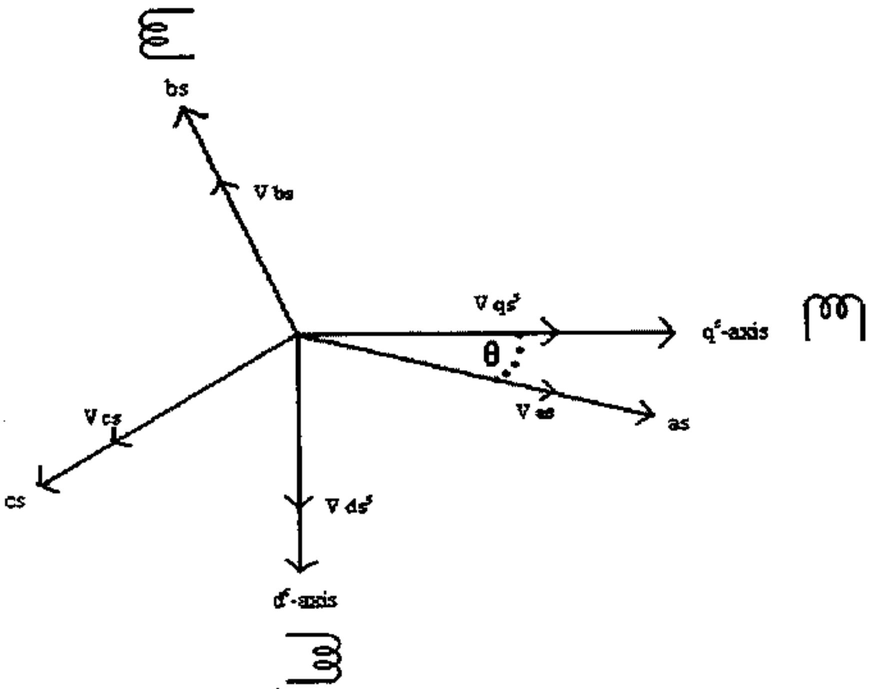


Fig.3.3 Stationary frame abc- ds-qs axes Transformation

Assume that the d^s - q^s axes are oriented at Θ angle, as shown in figure 3.1.

The voltages V_{ds}^s and V_{qs}^s can be resolved into as-bs-cs components and can be represented in the matrix form as

$$\begin{bmatrix} Vas \\ Vbs \\ Vcs \end{bmatrix} = \begin{bmatrix} \cos\theta & \sin\theta & 1 \\ \cos(\theta\text{-}120) & \sin(\theta\text{-}120) & 1 \\ \cos(\theta+120) & \sin(\theta+120) & 1 \end{bmatrix} \begin{bmatrix} Vqss \\ Vdss \\ Voss \end{bmatrix}$$

abc- dq Transformation

The corresponding inverse relation is,

$$\begin{bmatrix} Vqss \\ Vdss \\ Voss \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos\theta & \cos(\theta\text{-}120) & \cos(\theta+120) \\ \sin\theta & \sin(\theta\text{-}120) & \sin(\theta+120) \\ 0.5 & 0.5 & 0.5 \end{bmatrix} \begin{bmatrix} Vas \\ Vbs \\ Vcs \end{bmatrix}$$

dq - abc Transformation

where V_{0ss} is added as the zero sequence component, which may or may not be present. We have considered voltage as variable. The current and flux linkages can be transformed by similar equations.

It is convenient to set $\Theta = 0$, so that the q^s – axis is aligned with the as- axis. Ignoring the zero sequence components, the transformation can be simplified as,

$$\mathbf{V_{as} = V_{qs}^s} \quad (3.1)$$

$$\mathbf{V_{bs} = -.5\ V_{qs}^s - .866\ V_{ds}^s} \quad (3.2)$$

$$\mathbf{V_{cs} = -.5\ V_{qs}^s + .866\ V_{ds}^s} \quad (3.3)$$

And inversely

$$\mathbf{V_{qs}^s = (2/3)\ V_{as} - (1/3)\ V_{bs} - (1/3)\ V_{cs} = V_{as}} \quad (3.4)$$

$$\mathbf{V_{ds}^s = -(1/\sqrt{3})\ V_{bs} + (1/\sqrt{3})\ V_{cs}} \quad (3.5)$$

Figure 3.3 shows the synchronously rotating d^e- q^e axes, which rotate at synchronous speed ω_e with respect to the d^s- q^s axes and the angle $\Theta_e = \omega_e t$. The two phases d^s- q^s windings are transformed into the hypothetical windings mounted on the d^e- q^e axes. The voltages on the d^s- q^s axes can be converted into the d^e- q^e frame as follows,

$$\mathbf{V_{qs} = V_{qs}^s \cos\Theta_e - V_{ds}^s \sin\Theta_e} \quad (3.6)$$

$$\mathbf{V_{ds} = V_{ds}^s \sin\Theta_e + V_{qs}^s \cos\Theta_e} \quad (3.7)$$

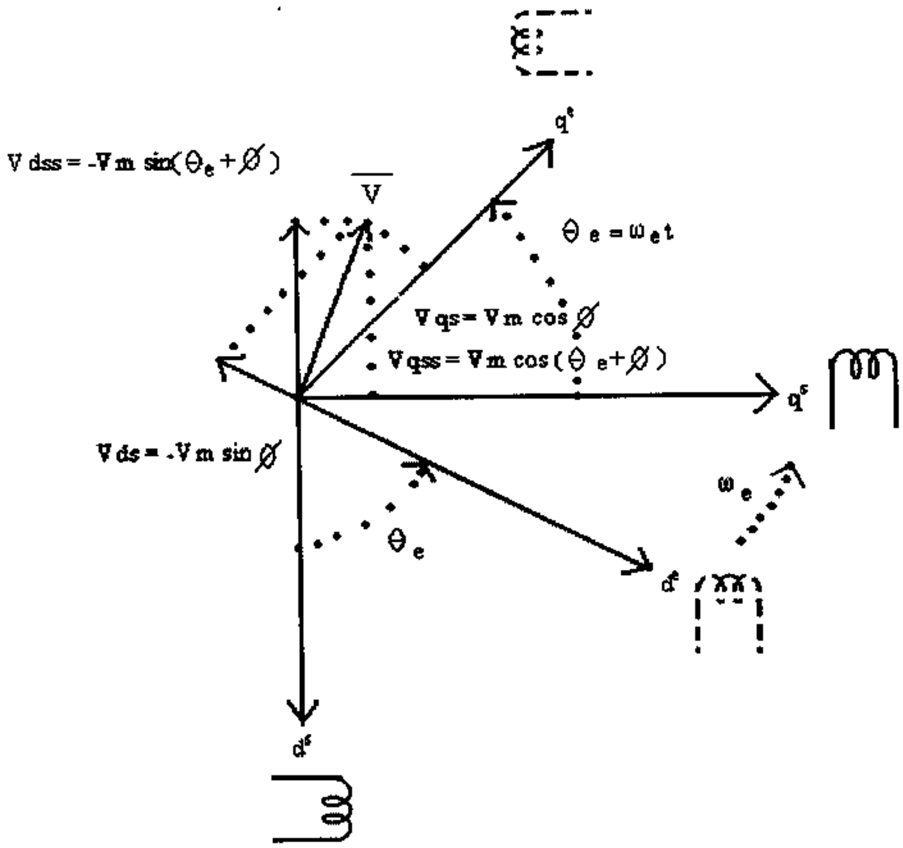


Fig 3.4 Stationary Frame to Synchronously Rotating Frame

For convenience, the subscript e has been dropped from now on the synchronously rotating frame parameters. Again resolving the rotating parameters into a stationary frame, the relation are

$$\mathbf{V_{qs}^{s} = V_{qs}\cos\Theta_e + V_{ds}\sin\Theta_e} \tag{3.8}$$

$$\mathbf{V_{ds}^{s} = -\,V_{qs}\sin\Theta_e + V_{ds}\cos\Theta_e} \tag{3.9}$$

As an example assume that the three phase stator voltage are sinusoidal and balanced, and are given by

$$V_{as} = V_m \cos(\omega et + \varphi) \quad (3.10)$$

$$V_{bs} = V_m \cos(\omega et - 2\pi/3 + \varphi) \quad (3.11)$$

$$V_{bs} = V_m \cos(\omega_e t + 2\pi/3 + \varphi) \quad (3.12)$$

Again substituting equations (3.10),(3.11),(3.12) in (3.4), (3.5) yields

$$V_{qs}^s = V_m \cos(\omega_e t + \varphi) \quad (3.13)$$

$$V_{ds}^s = -V_m \sin(\omega_e t + \varphi) \quad (3.14)$$

Again substituting (3.6),(3.7) in (3.13) , (3.14) ,we get

$$Vqs = Vm \cos\varphi \quad (3.15)$$

$$Vds = -Vm \sin\varphi \quad (3.16)$$

Equations (3.13), (3.14) show that V_{qs}^s and V_{ds}^s are balanced , two phase voltages of equal peak values and the latter is at $\pi/2$ angle phase lead with respect to the other component. Equation (3.15), (3.16) verify that the sinusoidal variables in a stationary frame appear as dc quantities in a synchronously rotating reference frame. This is an important derivation. Note that the stator variables are not necessarily balanced sinusoidal waves. They can be any arbitrary time functions.

3.3.2 Synchronously Rotating Reference Frame – Dynamic Model (Kron Equation)

For the two phase machine, we need to represent both $d^s - q^s$ and $d^r - q^r$ circuits and their variables in a synchronously rotating $d^e - q^e$ frame. We can write the following stator circuit equations:

$$V_{qs}^s = R_s\ i_{qs}^s + d/dt\,(\psi_{qs}^s) \quad (3.17)$$

$$V_{ds}^s = R_s\ i_{ds}^s + d/dt\,(\psi_{ds}^s) \quad (3.18)$$

where ψ_{qs}^s and ψ_{ds}^s are q-axis and d-axis stator flux linkages, respectively. When these equations are converted to $d^e - q^e$ frame , the following equations can be written,

$$V_{qs} = R_s\ i_{qs} + d/dt\ (\psi_{qs}) + \ \omega_e\ \psi_{ds} \tag{3.19}$$

$$V_{ds} = R_s\ i_{ds} + d/dt\ (\psi_{ds}) - \ \omega_e\ \psi_{qs} \tag{3.20}$$

Where all the variables are in rotating frame. The last term in the equations (3.19), (3.20) can be defined as speed emf due to rotation of these axes, that is , when $\omega_e = 0$, the equations revert to stationary frame. Note that the flux linkage in the d^e and q^e axes induce emf in the $d^e - q^e$ axes, respectively with $\pi/2$ lead angle.

If the rotor is not moving, that is $\omega_r = 0$, the rotor equation of a doubly fed wound rotor machine will be similar to equation (3.19), (3.20).

$$V_{qr} = R_r\ i_{qr} + d/dt\ (\psi_{qr}) + \ \omega_e\ \psi_{dr} \tag{3.21}$$

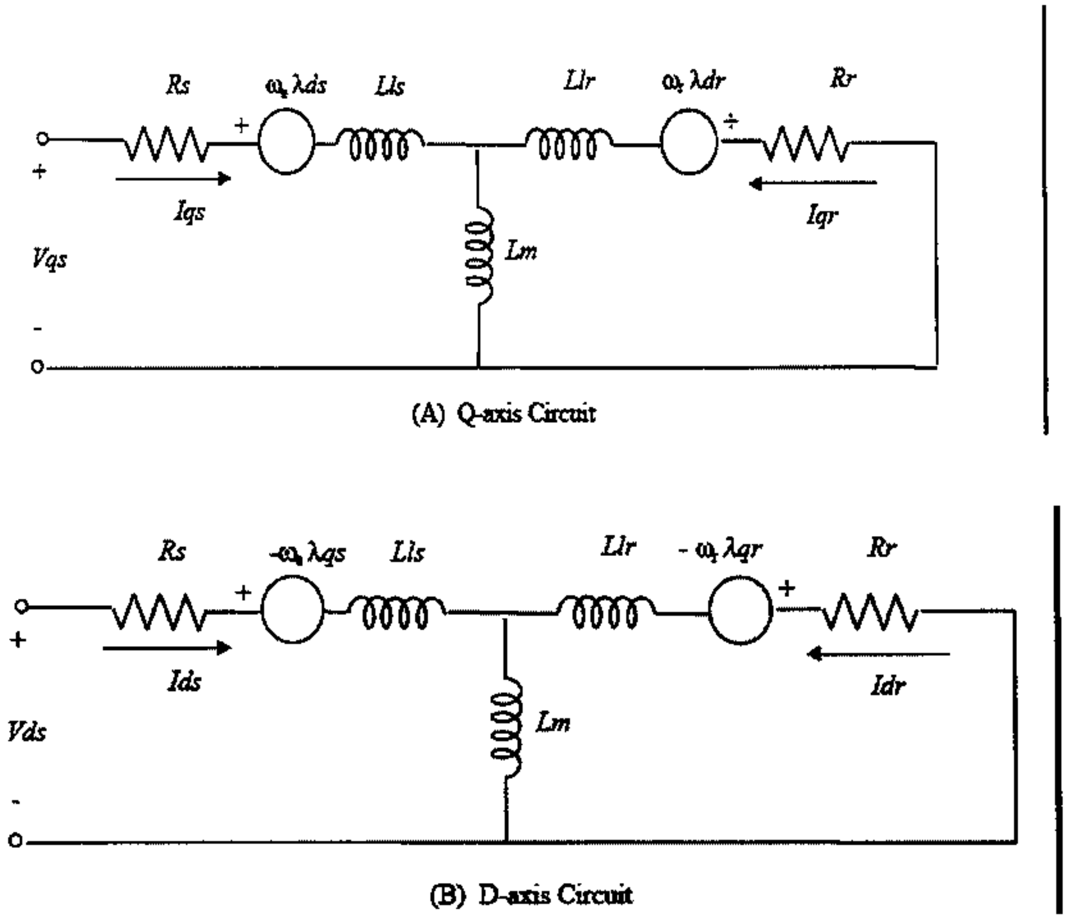
$$V_{dr} = R_r\ i_{dr} + d/dt\ (\psi_{dr}) - \ \omega_e\ \psi_{qr} \tag{3.22}$$

where all the variables and parameters are referred to the stator. Since the rotor actually moves at speed ω_r , the d –q axes fixed on the rotor move at a speed ω_e - ω_r relative to the synchronously rotating frame. Therefore, in the $d^e - q^e$ frame , the rotor equations should be modified as

$$V_{qr} = R_r\ i_{qr} + d/dt\ (\psi_{qr}) + \ (\ \omega_e - \omega_r\ )\ \psi_{dr} \tag{3.23}$$

$$V_{dr} = R_r\ i_{dr} + d/dt\ (\psi_{dr}) - (\ \omega_e - \omega_r\ )\ \psi_{qr} \tag{3.24}$$

The figure 3.5(A), 3.5(B) shows the $d^e - q^e$ dynamic model equivalent circuits that satisfy equations (3.19),(3.20)and (3.23),(3.24). A special advantage of the $d^e - q^e$ dynamic model of the machine is that all the sinusoidal variables in stationary frame appear as dc quantities in synchronous frame.



(A) Q-axis Circuit

(B) D-axis Circuit

Fig.3.5 d-q Equivalent circuit on a synchronous frame

The flux linkages expressions in terms of the currents can be written from figure 3.5 as follows

$\psi_{qs} = L_{is}\, i_{qs} + L_m\,(i_{qs} + i_{qr})$ (3.25)

$\psi_{qr} = L_{ir}\, i_{qs} + L_m\,(i_{qs} + i_{qr})$ (3.26)

$\psi_{qm} = Lm\,(i_{qs} + i_{qr})$ (3.27)

$\psi_{ds} = L_{is}\, i_{ds} + L_m\,(i_{ds} + i_{dr})$ (3.28)

$\psi_{qr} = L_{ir}\, i_{qr} + L_m\,(i_{ds} + i_{dr})$ (3.29)

$\psi_{dm} = L_m\,(i_{ds} + i_{dr})$ (3.30)

Combining the above expressions with equations (3.19),(3.20) and (3.23),(3.24) the electrical transient model in terms of voltages and currents can be given as

$$\begin{bmatrix} Vqs \\ Vds \\ Vqr \\ Vdr \end{bmatrix} = \begin{bmatrix} Rs + SLs & \omega eLs & SLm & eLm \\ -eLs & Rs + SLs & -\omega eLm & SLm \\ SLm & (\omega e - \omega r)Lm & Rs + SLr & (\omega e - \omega r)Lm \\ -(\omega e - \omega r)Lm & SLm & -(\omega e - \omega r)Lm & Rr + SLr \end{bmatrix} \begin{bmatrix} iqs \\ ids \\ iqr \\ idr \end{bmatrix}$$

Where S is the Laplace operator. For a singly fed machine, such as a cage motor, $V_{qr} = V_{dr} = 0$.

If the speed ω_r is considered (infinite inertia load), the electrical dynamics of the machine are given by a fourth order linear system. Then by knowing the inputs V_{qs},V_{ds} and ω_e, the currents i_{qs} , i_{ds} , i_{qr} and i_{dr} can be solved from MATRIX equation. If the machine is fed by current source i_{ds} , i_{qs} and ω_e are independent. Then the dependent variables V_{qs} , V_{ds}, i_{qr} and i_{dr} can be solved from MATRIX equation.

The speed ω_r in equation MATRIX cannot normally be treated as a constant. It can be related to the torque as

$$\mathbf{T_e = T_l + J\ d/dt\ (\omega_m) = T_l + 2/p\ J\ d/dt\ (\omega_r)} \qquad (3.31)$$

Where T_l = load torque ,

J = rotor inertia ,

ω_m = mechanical speed.

3.3.3 Stationary Frame – Dynamic Model (Stanley Equation)

The dynamic machine model in stationary frame can be derived simply by substituting $\omega_e = 0$ in equation (3.19),(3.20) and (3.23),(3.24) . The corresponding frame equations are given as

$$\mathbf{V_{qs}^s = R_s\ i_{qs}^s + d/dt\ (\psi_{qs}^s)} \qquad (3.32)$$

$$\mathbf{V_{ds}^s = R_s\ i_{ds}^s + d/dt\ (\psi_{ds}^s)} \qquad (3.33)$$

$$\mathbf{R_r\ i_{qr}^s + d/dt\ (\psi_{qr}^s) + \omega_e\ \psi_{dr}^s = 0} \qquad (3.34)$$

$$\mathbf{R_r\ i_{dr}^s + d/dt\ (\psi_{dr}^s) - \omega_e\ \psi_{qr}^s = 0} \qquad (3.35)$$

Where $V_{qr} = V_{dr} = 0$. Figure 3.7 shows the corresponding equivalent circuits. As mentioned before, in the stationary frame, the variables appear as sine waves in steady state with sinusoidal inputs.

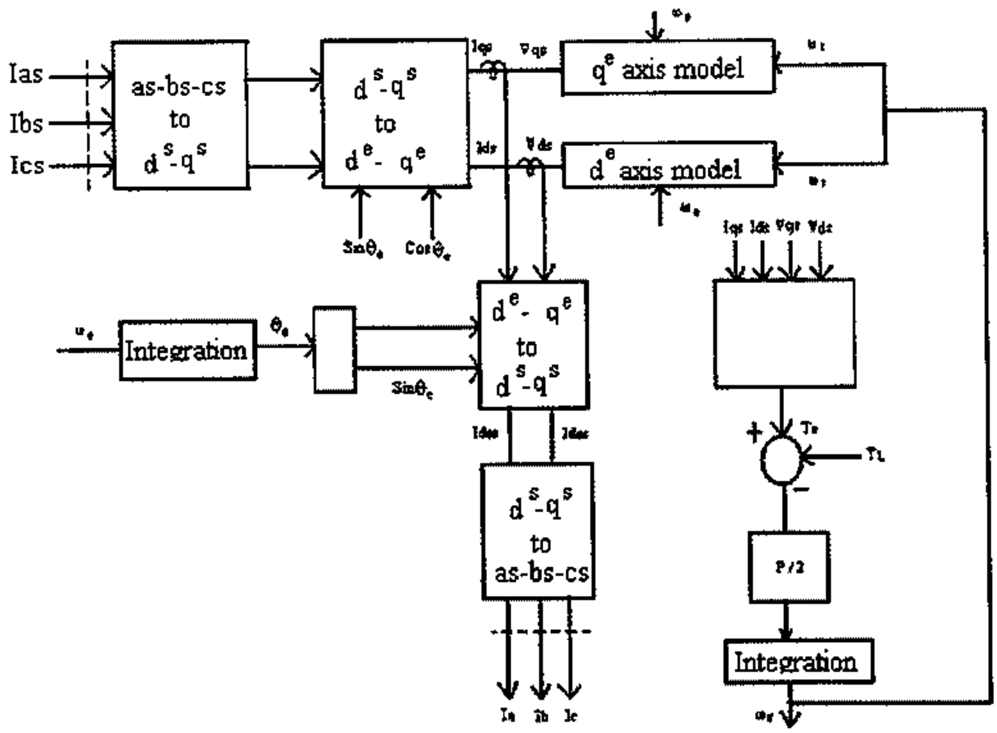


Fig.3.6. Synchronously Rotating Frame Machine Model

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

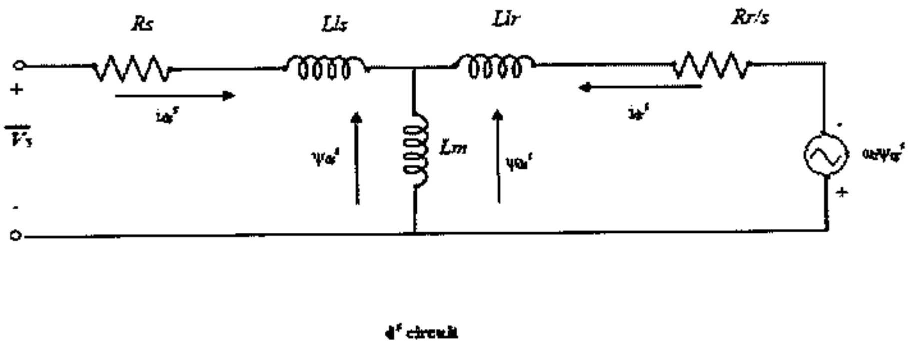
$$T_e = 1.5\,(p/2)\,(\psi_{dm}^s\, i_{qr}^s - \psi_{qm}^s\, i_{dr}^s) \quad (3.36)$$

$$T_e = 1.5\,(p/2)\,(\psi_{dm}^s\, i_{qs}^s - \psi_{qm}^s\, i_{ds}^s) \quad (3.37)$$

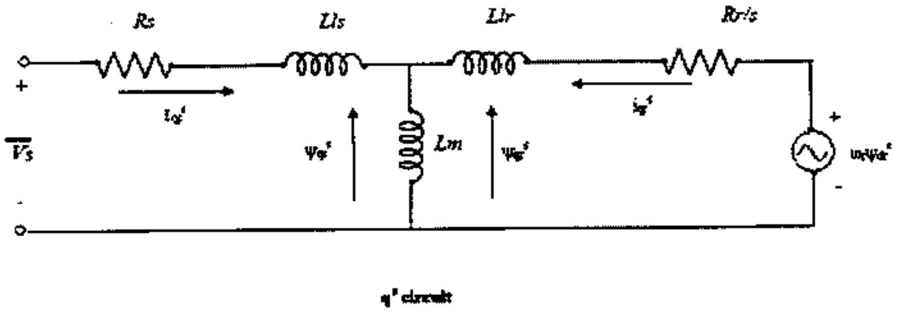
$$T_e = 1.5\,(p/2)\,(\psi_{ds}^s\, i_{qs}^s - \psi_{qs}^s\, i_{ds}^s) \quad (3.38)$$

$$T_e = 1.5\,(p/2)\,(i_{qr}^s\, i_{qs}^s - i_{qr}^s\, i_{ds}^s) \quad (3.39)$$

$$T_e = 1.5\,(p/2)\,(\psi_{dr}^s\, i_{qr}^s - \psi_{qr}^s\, i_{d1r}^s) \quad (3.40)$$



(a)



(b)

Fig.3.7. $d^s - q^s$ equivalent circuit

Equations 3.32, 3.33, 3.34 and 3.35 can be easily combined to derive complex model as

$$\mathbf{V_{qds}^s = R_s\ i_{qds}^s + d/dt\,(\psi_{qds}^s)} \quad (3.41)$$

$$\mathbf{R_r\ i_{qdr}^s + d/dt\,(\psi_{qdr}^s) - j\,\omega_r\,\psi_{qdr}^s = 0} \quad (3.42)$$

Where $V_{qds}^s = V_{qs}^s - j\ V_{ds}^s$, $\psi_{qds}^s = \psi_{qs}^s - j\ \psi_{ds}^s$, $i_{qds}^s = i_{qs}^s - j\ i_{ds}^s$ etc. The complex equivalent circuit in stationary frame is shown in figure 3.8(a), 3.8(b).

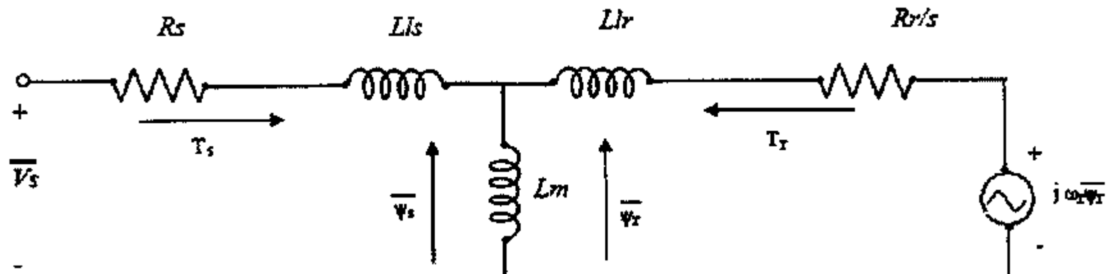


Fig.3.8 (a) Stationary frame complex equivalent circuit

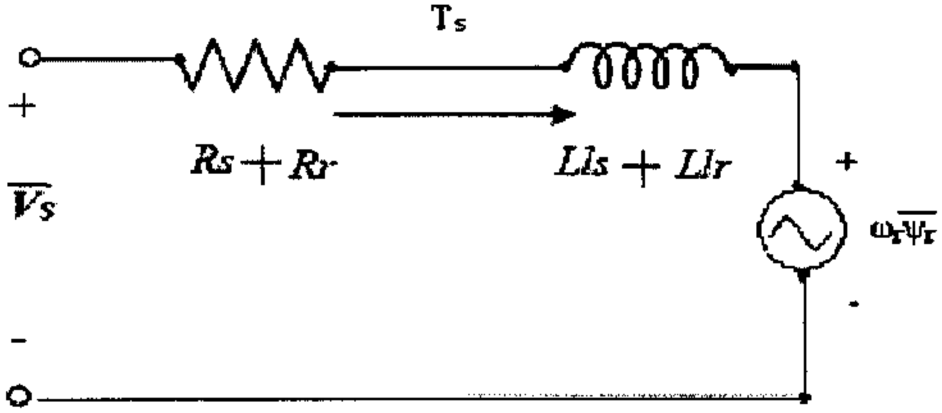


Fig.3.8 (b) Simplified per phase equivalent circuit

3.4 Direct Vector Control

The basic block diagram of the direct vector control method for a PWM voltage fed inverter drive is shown in Fig 2.9. The principal vector control parameters, i^*_{ds} and i^*_{qs}, which are dc values in synchronously rotating frame, are converted into stationary frame (defined as vector rotation (VR)) with the help of a unit vector ($\cos\theta_e$ and $\sin\theta_e$) generated from flux vector signals ψ_{dr}^s and ψ_{qr}^s. The resulting stationary frame signals are then converted to phase current commands for the inverter. The flux signals ψ_{dr}^s and ψ_{qr}^s are generated from the machine terminal voltages and currents with the help of the voltage model estimator. A flux control loop has been added for precision control of flux. The torque component of current i^*_{qs} is generated from the speed control loop through a bipolar limiter. The torque, proportional to I_{qs} (with constant flux), can be bipolar. It is negative with negative I_{qs}, and correspondingly, the phase position of I_{qs} becomes negative. An additional torque control loop can be added within the speed loop, if desired. Fig 2.9 can be extended to field weakening mode by programming the flux command as a function of speed so that the inverter remains in PWM mode. Vector control by current regulation is lost if the inverter attains the square-wave mode of operation.

The correct alignment of current I_{ds} in the direction of flux ψ_r and the current I_{qs} perpendicular to it are crucial in vector control. This alignment, with the help of stationary frame rotor flux vectors ψ_{dr}^s and ψ_{qr}^s. The $d^e - q^e$ frame is rotating at synchronous speed ω_e with respect to stationary frame - q^s, and at any instant, the angular position of the d^e axis with respect to the d^s axis is θ_e, where $\theta_e = \omega_e t$. so we can write the following equations:

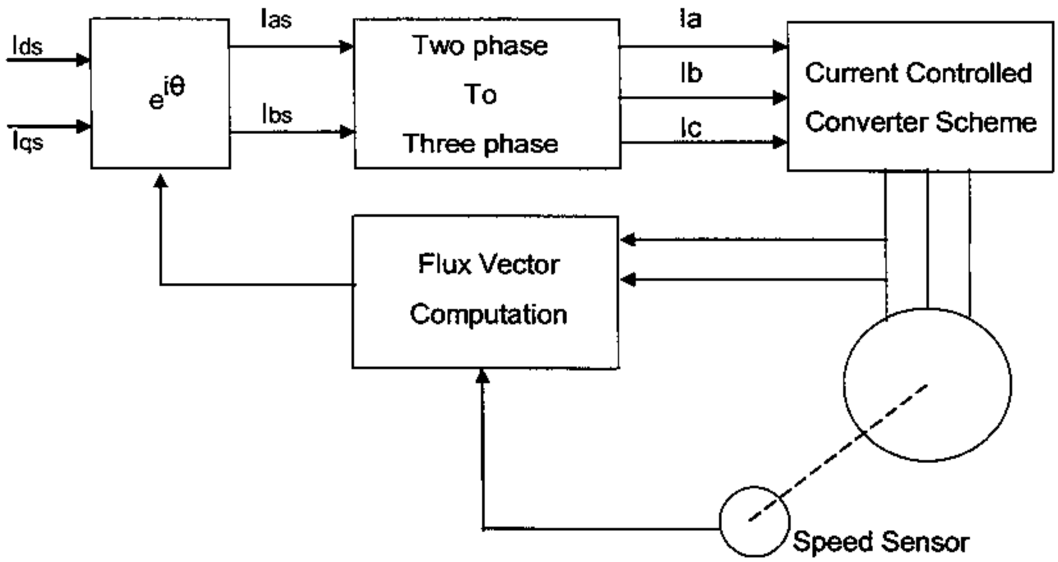


Fig 3.9 Direct Vector Control of Induction Motor

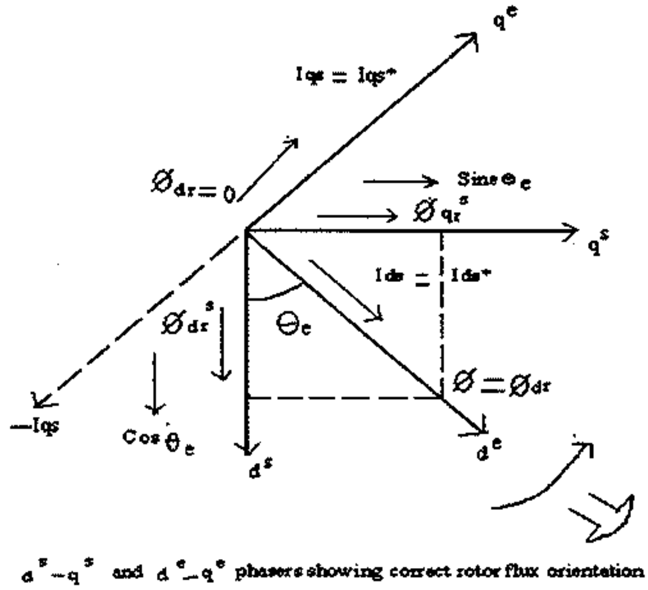


Figure 3.10 Phasor diagram of direct vector control

$$\psi_{dr}^{s} = \psi_r \cos \theta_e \quad (3.43)$$

$$\psi_{qr}^{s} = \psi_r \sin \theta_e \quad (3.44)$$

In other words,

$$\cos \theta_e = \psi_{dr}^{s} / \psi_r \quad (3.45)$$

$$\sin \theta_e = \psi_{qr}^{s} / \psi_r \quad (3.46)$$

$$\psi_r = \sqrt{(\psi_{dr}^{s} + \psi_{qr}^{s})} \quad (3.47)$$

Where vector ψ_r is represented by magnitude ψ_r. Signals cos θ_e and sin θ_e have been plotted in correct phase position in Fig 3.10. These unit vector signals, when used for vector rotation in Fig 3.10, give a ride of current I_{ds} on the d^e-axis (direction of ψ_r) and current I_{qs} on the q^e-axis. At this condition, $\psi_{dr} = \psi_r$ and $\psi_{qr} = 0$, and the corresponding torque expression is given by the equation (3.6) like a dc machine. When the I_{qs} polarity is reversed by the speed loop, the I_{qs} position also reverses, giving negative torque. The generation of a unit vector signal from feedback flux vectors gives the name "DIRECT VECTOR CONTROL".

4. INDIRECT VECTOR CONTROL

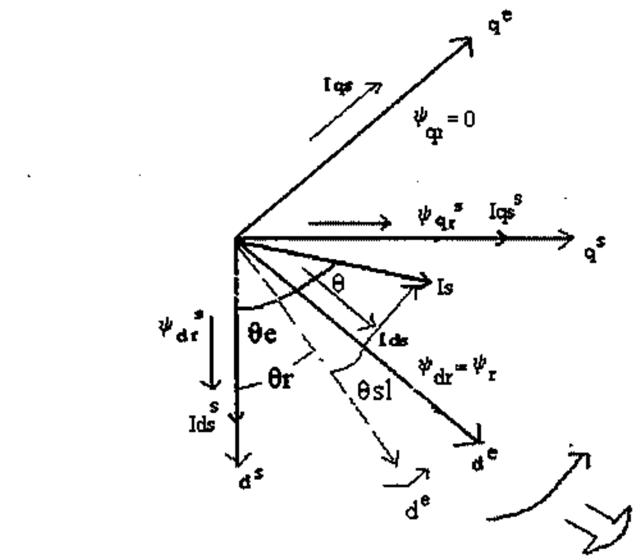
4.1. BASIC PRINCIPLE

The indirect vector control method is essentially the same as the direct vector control, except the unit vector signals (cos θ_e and sin θ_e) are generated in feed forward manner. Indirect vector control is very popular in industrial applications. Fig 4.1 explains the fundamental principle of feed forward vector control with the help of a phasor diagram. The $d^s - q^s$ axes, which are fixed on the rotor, are moving at speed ω_r as shown in Fig 4.1. Synchronously rotating axes $d^e - q^e$ are rotating ahead of the $d^r - q^r$ axes by the positive slip angle θ_{sl} corresponding to slip frequency ω_{sl}. Since the rotor pole is directed on the d^e axis and $\omega_e = \omega_r + \omega_{sl}$, we can write

$$\boldsymbol{\theta_e = \int \omega_e\, dt = \int (\omega_r + \omega_{sl})\, dt = \theta_r + \theta_{sl}} \quad \textbf{(4.1)}$$

Note that the rotor pole position is not absolute, but is slipping with respect to the rotor at frequency ω_{sl}. The phasor diagram suggests that for decoupling control, the stator flux component of current I_{ds} should be aligned on d^e axis, and the torque component of current I_{qs} should be on the q^e axis is shown in Fig 4.1.

For decoupling control, we can now make a derivation of control equations of indirect vector control with the help of $d^e - q^e$ equivalent circuits. The rotor circuit equations can be written as



$d^s - q^s$ and $d^e - q^e$ phasers showing correct rotor flux orientation

Figure 4.1 Phasor diagram explaining vector control

$$\mathbf{d/dt\ (\psi_{dr}) + Rr\ i_{dr} - (\omega_e - \omega_r)\ \psi_{qr} = 0} \quad (4.2)$$

$$\mathbf{d/dt\ (\psi_{qr}) + Rr\ i_{qr} - (\omega_e - \omega_r)\ \psi d_r = 0} \quad (4.3)$$

The rotor flux linkage expressions can be given as

$$\mathbf{\psi_{dr} = L_r\, i_{dr} + L_m\, i_{ds}} \quad (4.4)$$

$$\mathbf{\psi_{qr} = L_r\, i_{qr} + L_m\, i_{qs}} \quad (4.5)$$

From the above equations, we can write

$$\mathbf{I_{dr} = (1/Lr)\ \psi_{dr} - (L_m/L_r)\ i_{ds}} \quad (4.6)$$

$$\mathbf{I_{qr} = (1/Lr)\ \psi_{qr} - (L_m/L_r)\ i_{qs}} \quad (4.7)$$

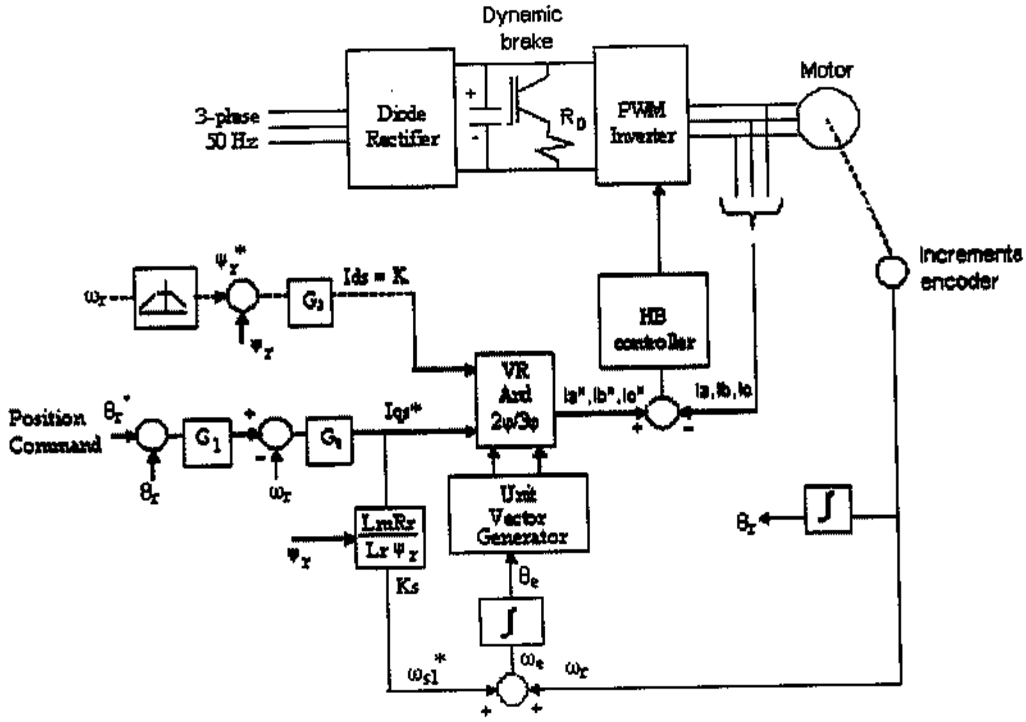


Figure 4.2 Indirect vector control with open loop flux control

The rotor currents in Equations (4.2) and (4.3), which are inaccessible, can be eliminated with the help of equations (4.6) and (4.7) as

$$\mathbf{d/dt\ (\psi_{dr}) + (R_r/L_r)\ \psi_{dr} - (L_m/L_r)\ R_r\ i_{ds} - \omega_{sl}\ \psi_{qr} = 0} \quad (4.8)$$

$$\mathbf{d/dt\ (\psi_{qr}) + (R_r/L_r)\ \psi_{qr} - (L_m/L_r)\ R_r\ i_{qs} - \omega_{sl}\ \psi_{dr} = 0} \quad (4.9)$$

Where $\omega_{sl} = \omega_e - \omega_r$ has been substituted.

For decoupling control, it is desirable that

$$\mathbf{\psi_{qr} = 0} \quad (4.10)$$

that is,

$$\mathbf{d/dt\ (\psi_{qr}) = 0} \tag{4.11}$$

so that the total rotor flux ψ_r is directed on the d^e axis.

Substituting the above conditions in Eqns (4.8) and (4.9), we get

$$\mathbf{(L_r/R_r)\ d/dt\ (\psi_r) + \psi_r = L_m\ i_{ds}} \tag{4.12}$$

$$\mathbf{\omega_{sl} = (L_m\ R_r\ i_{qs})/(\ \psi_r L_r)} \tag{4.13}$$

where $\psi_r = \psi_{dr}$ has been substituted.

If rotor flux ψ_r = constant, which is usually the case, then from Eqn (3.23),

$$\mathbf{\psi_r = L_m\ i_{ds}} \tag{4.14}$$

In other words, the rotor flux is directly proportional to current i_{ds} in steady state.

To implement the indirect vector control strategy, it is necessary to take Equations (4.1), (4.12) and (4.13) into consideration. Fig 4.2 shows a Induction motor using the indirect vector control method. The power circuit consists of a front-end diode rectifier and a PWM inverter with a hysteresis-band current control is used. The speed controller loop generates the torque component of current i^*_{qs}. The flux component of current i^*_{ds} for the desired rotor flux ψ_r is determined from equation (4.14), and is maintained constant here in the open loop manner for simplicity. The variation of magnetising inductance L_m will cause some drift in the flux. The slip frequency ω_{sl}^* is generated from i^*_{qs} in feed forward manner from Equation (4.13) to satisfy the phasor diagram in Fig 4.1. the corresponding expression of slip gain K_s is given as

$$\mathbf{K_s = \omega_{sl}^*/i_{qs}^* = (L_m R_r)/(L_r \psi_r)} \tag{4.15}$$

Signal ω_{sl}^* is added with speed signal ω_r to generate frequency signal ω_e. The unit vector signals cos θ_e and sin θ_e are than generated from ω_e by integration and look-up tables. The VR and $2\varphi/3\varphi$ transformation are the same as in Fig 3.9. The speed signal from an encoder is mandatory in indirect vector control because the slip signal locates the pole with respect to the rotor d^r axis in feed forward manner, which is moving at speed ω_r. an absolute pole position on the rotor is not required in this case like a synchronous motor. If the polarity of i^*_{qs} becomes negative for negative torque, the phasor i_{qs} in Fig 3.2 will be reversed, and correspondingly, ω_{sl} will be negative (i.e., θ_{sl} is negative), which will shift the rotor pole position (d^e axis) below the d^r axis. The speed control range in indirect vector control can easily be extended from stand-still (zero speed) to the field-weakening region. The addition of field-weakening control is shown by the dotted block diagram and the corresponding operation explained in Fig 4.3 In this case, close loop flux control is necessary. In the constant torque region, the flux is constant. However, in the field- weakening region, the flux is programmed such that the inverter always operates in PWM mode. The loss of torque and power for field-weakening vector control are indicated in the fig.4.3.

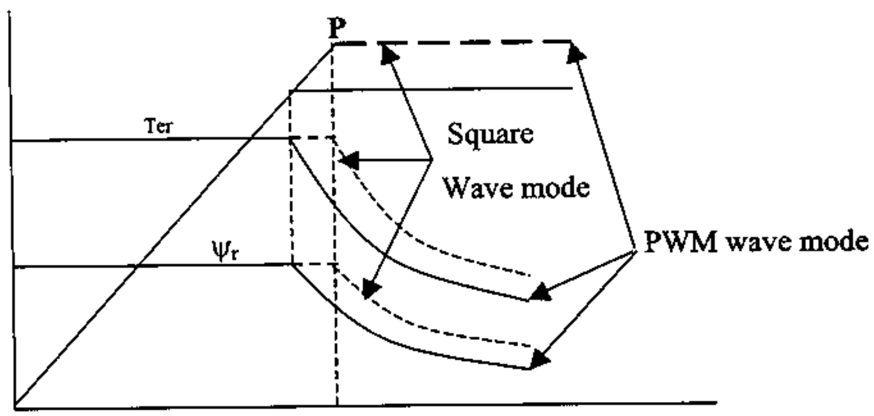


Figure 4.3 Torque- speed curves including field- weakening mode

In the indirect vector control method, instantaneous current control of inverter is necessary. For that Hysteresis- band PWM current control can be used.

4.2 VOLTAGE SOURCE INVERTER

Voltage Source Inverters, as the name indicates, receive dc voltage at one side and convert to ac voltage on the other side. The ac voltage and frequency may be variable or constant depending on the application. So the inverter is needed to give required ac voltage to the system. The inverter output can be single-phase or polyphase and can have square wave, sine wave, PWM wave, stepped wave or quasi-square wave at the output. Here we use Insulated Gate Bipolar Transistor (IGBT) Inverter. Because the IGBT has the advantage of Bipolar Junction Transistor (Switching loss is minimum) and also has the advantage of Metal Oxide Semiconductor Field Effective Transistor (MOSFET) like Input impedance is high. We use the PWM Technique to give pulses to IGBT.

4.3. PULSE WIDTH MODULATION

An inverter contains electronic switches, it is possible to control the output voltage as well as optimize the harmonics by performing multiple switching within the inverter with the constant dc input voltage V_d.

PWM classification

The classifications of PWM techniques can be given as follows:

- Sinusoidal PWM (SPWM)
- Selected harmonic elimination (SHE) PWM
- Minimum ripple current PWM
- Space-vector PWM (SVM)
- Random PWM
- Hysteresis band current control PWM
- Sinusoidal PWM with instantaneous current control
- Delta modulation
- Sigma-delta modulation

The hysteresis band PWM has been chosen in our project because of its simple implementation, fast transient response, direct limiting of device peak current, and practical insensitivity of dc link voltage ripple that permits a lower filter capacitor.

4.4. HYSTERESIS BAND CURRENT CONTROL PWM:

Hysteresis-band PWM is basically an instantaneous feedback current control method of PWM where the actual current continually tracks the command current within a hysteresis band. Figure 4.4 explains the operation of hysteresis-band PWM current control. The control circuit generates the sine reference current wave of desired magnitude and frequency, and it is compared with the actual phase current wave. As the current exceeds a prescribed hysteresis band, the upper switch in the half-bridge is turned off and the lower switch is turned on. As a result, the output voltage transitions from $+0.5V_d$ to $-0.5\ V_d$, and the current starts to decay. As the current crosses the lower band limit, the lower switch is turned off and the upper switch is turned on. Lock out time t_d is provided at each transition to prevent a shoot through fault. The actual current wave is thus forced to track the sine reference wave within the hysteresis band by back- and-forth switching of the upper and lower switches. The inverter then essentially becomes a current source with peak to peak current ripple, which is controlled within the hysteresis band irrespective of V_d fluctuation. When the upper switch is closed, the positive slope the current is given as

$di/dt = (0.5Vd - V_{cm} \sin\omega_e t) / L$ (3.16)

Where 0.5 V_d is the applied voltage, V_{cm} sinωt=instantaneous value of the opposing load CEMF, and L=effective load inductance. The corresponding equation when the lower switch is closed is given as

$di/dt = - (0.5V_d + V_{cm} \sin\omega_e t) / L$ (3.17)

The peak to peak current ripple and switching frequency are related to the width of the hysteresis band. For example, a smaller band will increase switching frequency and lower the ripple. An optimum band that maintains a balance between the harmonic ripple and inverter switching loss is desirable. The hysteresis band PWM can be smoothly transitioned to square wave voltage mode through the quasi-PWM region. In the lower-speed region of the machine when the CEMF is low, there is no difficulty in the current controller tracking. However at higher speeds , the current controller will saturate in the part of the cycle due to a higher CEMF and fundamental frequency-related harmonics

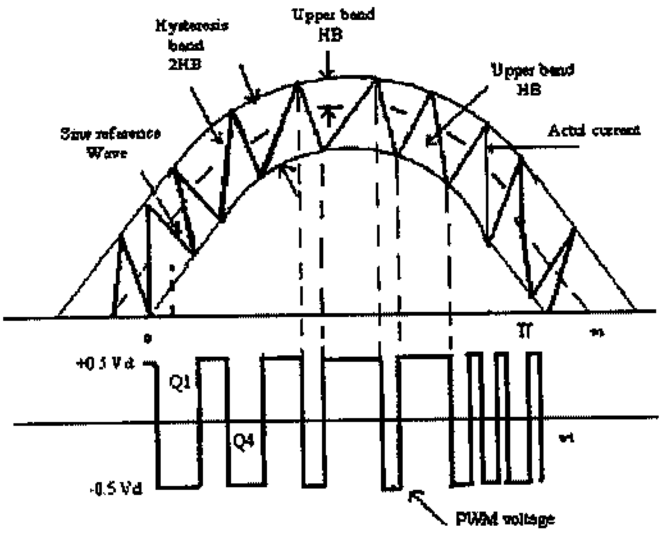


Fig.4.4 Hysteresis band control

will appear. At this condition, the fundamental current will be less and its phase will lag with respect to the command current.

5. FUZZY LOGIC CONTROLLER

5.1 INTRODUCTION

The concept of Fuzzy Logic (FL) was conceived by Lotfi Zadeh, a professor at the University of California at Berkley, and presented as a way of processing data by allowing partial set membership rather than crisp set membership or non-membership. This approach to set theory was not applied to control systems until the 70's due to insufficient small-computer capability prior to that time. Professor Zadeh reasoned that people do not require precise, numerical information input, and yet they are capable of highly adaptive control. If feedback controllers could be programmed to accept noisy, imprecise input, they would be much more effective and perhaps easier to implement.

FL is a problem-solving control system methodology that lends itself to implementation in systems ranging from simple, small, embedded micro-controllers to large, networked, multi-channel PC or workstation-based data acquisition and control systems. It can be implemented in hardware, software, or a combination of both. FL provides a simple way to arrive at a definite conclusion based upon vague, ambiguous, imprecise, noisy, or missing input information. FL's approach to control problems mimics how a person would make decisions, only much faster.

The concept of crisp sets comes from traditional or classical logic. Crisp sets have rigid membership requirements where every object is either completely included or excluded from a set. In contrast to this true or false scenario, fuzzy sets allow for continuous-set membership values ranging from 0 to 1.

5.2. FUZZY SET

In traditional logic, sets are "crisp", either you belong 100% to a set or you do not. A set of tall people may consist of all people over six feet tall, anyone less than six feet is "short" (or more appropriately "not tall"). Fuzzy logic allows sets to be "fuzzy" so anyone over six feet tall may have 100% membership in the "tall" set, but may also have 20% membership in the "medium height" set. Also referred to as: - fuzzy subset.

5.2.1. Membership Functions

The membership function is a graphical representation of the magnitude of participation of each input. It defines a fuzzy set by associating every element in the set with a number between 0 and 1.

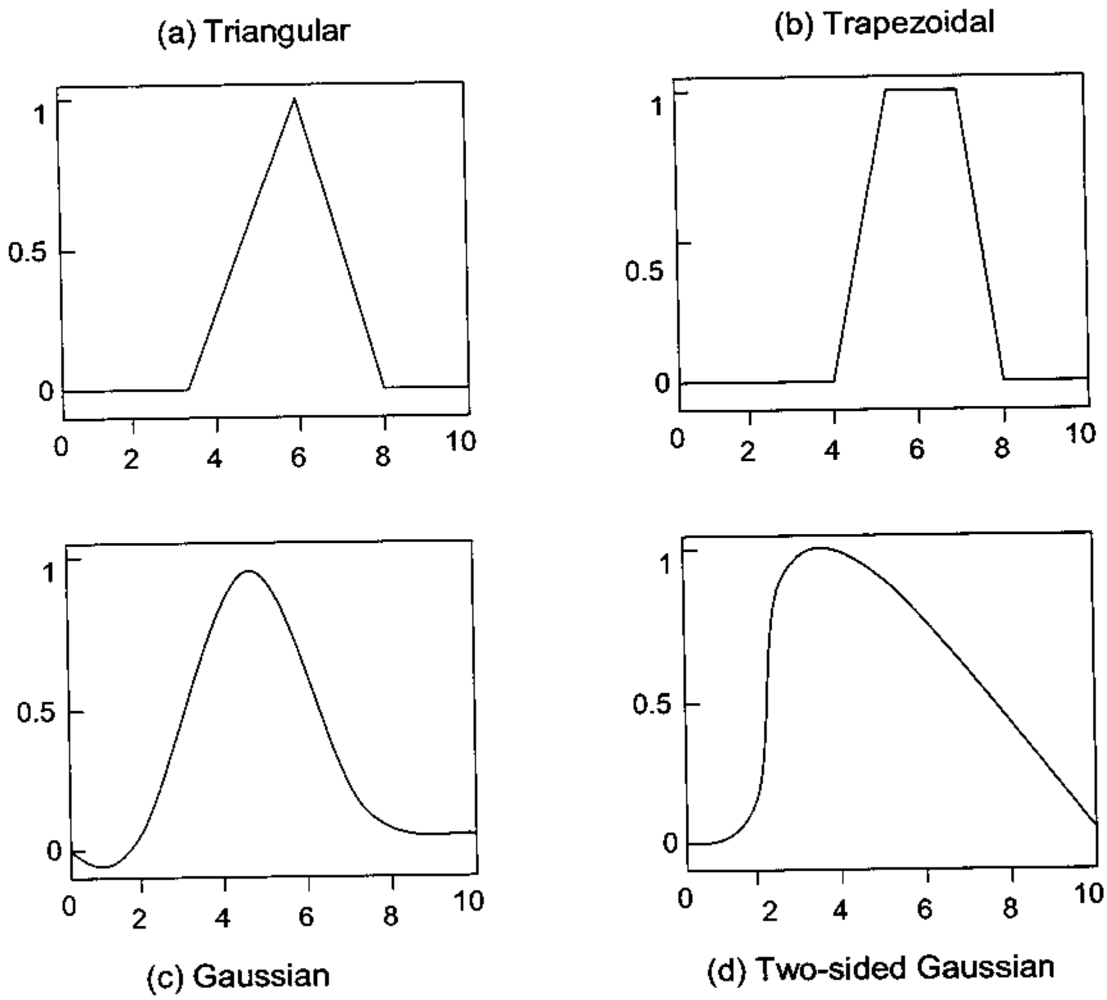


Fig 5.1. Different types of membership functions

The simplest and most commonly used MF is the triangular type, which can be symmetrical or unsymmetrical in shape. A trapezoidal MF (symmetrical or unsymmetrical) has the shape of truncated triangle. The other types of MFs are Gaussian, bell, Two-sided Gaussian etc. In practice the triangular and Gaussian are more enough to solve problems.

5.2.2. Fuzzy Implication Methods

The fuzzy implication is the process of shaping the fuzzy set in the consequent part based on the antecedent part of a role. The different types of implication methods are

1. Mamdani method
2. Lusing Larson Type
3. Sugeno Type

The most commonly used implication type is mamdani method. In the mamdani method based on the inputs and membership function the output is calculated. In the sugeno method the output MFs are constants or have linear relations with the inputs.

5.2.3. Defuzzification

This is the process of transforming a fuzzy output of a fuzzy system into a crisp output. The important methods of defuzzification are

5.2.3.1. Center of Area (COA) Method

In the COA method (often called the center of gravity method) of defuzzification, the crisp output Z_0 of the Z variable is taken to be the geometric center of the output fuzzy value μ_{out} (Z) area, where μ_{out} (Z) is formed by taking the union of all the contributions of rules whose DOF>0. The general expression for COA defuzzification is

$$Z_0 = \frac{\sum_{i=1}^{n} Zi\ \mu out\ (Zi)}{\sum_{i=1}^{n} \mu out\ (Zi)} \qquad (4.1)$$

COA defuzzification is a well- known method and it is often used in spite of some amount of complexity in the calculation. Note that if the areas of two or more contributing rules overlap, the overlapping area is counted only once.

5.2.3.2. Height Method

In the height method of defuzzification, the COA method is simplified to consider only the height of each contributing MF at the mid- point of the base. The accuracy of the output is slightly less than COA method.

5.2.3.3. Mean of Maxima (MOM) Method

The height method of defuzzification is further simplified in the MOM method, where only the highest membership function component in the output is considered. If M such maxima are present, then the formula is

$$Z_0 = \sum_{m=1}^{M} \frac{Zm}{M} \qquad (4.2)$$

Where Z_m = m^{th} element in the universe of discourse, where the output MF is at the maximum value and M = number of such elements.

5.2.3.4. Sugeno Method

In the sugeno method, the defuzzification formula is

$$Z_0 = \frac{K1.\,DOF1 + K2.\,DOF2 + K3.\,DOF3}{DOF1 + DOF2 + DOF3} \qquad (4.3)$$

Whereas in the first- order method, the formula is

$$Z_0 = \frac{Z1.DOF1 + Z2.DOF2 + Z3.\,DOF3}{DOF1 + DOF2 + DOF3} \qquad (4.4)$$

5.2 CONTROLLER DESIGN

The Speed controller in the ordinary Field oriented control method is based on the design of the normal PI controller design. But the ordinary PI controller is affected by the variations in the machine parameters. So we use the intelligent control methods for the speed controller design. The Fuzzy logic based controller doesn't affected by any machine parameters. So the robustness of the system is improved. Here the fuzzy logic controller is used to control the speed of the motor. The structure of the Fuzzy Logic Controller is shown in Fig. and the Fuzzy logic Rules used in the controller is shown in table 1.

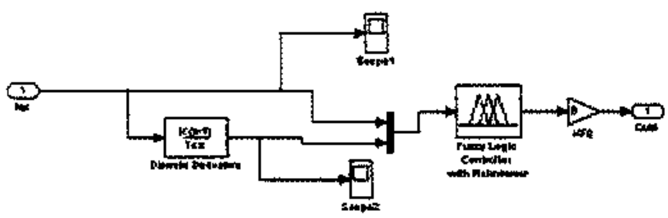


Figure 5.2. Fuzzy Logic Controller

The FLC rules are based on

Ri: IF Ai AND Bi THEN Ci

The controller structure is Proportional- Integral (PI) and illustrated in fig.5.2.

Here **Kp = kuR (KIqs) and Ki = KuR (KIqs*)** are the controller proportional and integral gains and R (.) is defined by the controller rule base which is summarised in table.1. Here NB (Negative Big), NM (Negative Medium), NS (Negative Small), Z (Zero), PS (positive Medium) and PB (Positive Big) are linguistic variables. The max-min inference method (tests the magnitudes of each rule and selects the highest one) was used and the defuzzification procedure was based on the centre of area method (The horizontal coordinate of the "fuzzy centroid" of the area under that function is taken as the output).

Table 1. Fuzzy Logic Controller Rule

Iqs*

Iqs

	NB	NM	NS	Z	PS	PM	PB
NB	NVB	NVB	NB	NM	NS	NS	Z
NM	NVB	NB	NM	NS	NS	Z	PS
NS	NB	NM	NS	NS	Z	PS	PS
Z	NM	NS	NS	Z	PS	PS	PM
PS	NS	NS	Z	PS	PS	PM	PB
PM	NS	Z	PS	PS	PM	PB	PVB
PS	Z	PS	PS	PM	PB	PVB	PVB

4.3 ADVANTAGES OF FUZZY LOGIC

- Fuzzy logic is conceptually easy to understand.

The mathematical concepts behind fuzzy reasoning are very simple. What makes fuzzy nice is the "naturalness" of its approach and not its far-reaching complexity.

- Fuzzy logic is flexible.

With any given system, it's easy to massage it or layer more functionality on top of it without starting again from scratch.

- Fuzzy logic is tolerant of imprecise data.

Everything is imprecise if you look closely enough, but more than that, most things are imprecise even on careful inspection. Fuzzy reasoning builds this understanding into the process rather than tacking it onto the end.

- Fuzzy logic can model nonlinear functions of arbitrary complexity.

You can create a fuzzy system to match any set of input-output data. This process is made particularly easy by adaptive techniques like ANFIS (Adaptive Neuro-Fuzzy Inference Systems), which are available in the Fuzzy Logic Toolbox.

- Fuzzy logic can be built on top of the experience of experts.

In direct contrast to neural networks, which take training data and generate opaque, impenetrable models, fuzzy logic lets you rely on the experience of people who already understand your system.

- Fuzzy logic can be blended with conventional control techniques.

Fuzzy systems don't necessarily replace conventional control methods. In many cases fuzzy systems augment them and simplify their implementation.

- Fuzzy logic is based on natural language.

The basis for fuzzy logic is the basis for human communication. This observation underpins many of the other statements about fuzzy logic.

6. ADAPTIVE CONTROL

6.1 INTRODUCTION

A linear control system with invariant plant parameters can be designed easily with the classic design techniques, such as Nyquist and Bode plots. Ideally, a vector-controlled ac drive can be considered as linear, like a dc drive system. However, in industrial and mechanical parameters of the drive hardly remain constant. Besides, there is a load torque disturbance effect. For example, the inertia of an electric vehicle or subway drive will vary with passenger load. In a robotic drive, on the other hand, the inertia will change, depending on the length of the arm and the load it carries. In a rolling mill drive, the drive load torque will change abruptly when a metal slab is introduced within the rolls. Figure 6.1 shows a block diagram of a speed-controlled vector drive indicating the moment of inertia J and load torque T_L variation. For the fixed control parameters in G, an increase of the J parameter will reduce the loop gain, deteriorating the system's performance. Similarly, a sudden increase of load torque T_L or J will temporarily reduce the speed until it is compensated by the sluggish speed loop. The effect of the parameter variation can be compensated to some extent by a high-gain negative speed back loop. But, excessive gain may cause an underdamping or instability problem in extreme cases. The problem discussed above requires adaptation of the controller G in real time, depending on the plant variation and load torque disturbance, so that the system response is not affected. Basically different types of adaptive control methods are used. Adaptive control based on artificial intelligence is also used in many applications.

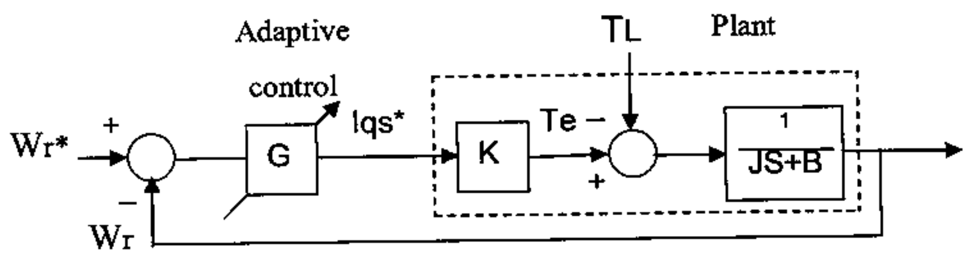


Figure 6.1 Adaptive control block diagram

Adaptive control techniques can be generally classified as

- Self-tuning control
- MRAC
- Sliding mode or variable structure control
- Expert system control
- Fuzzy control
- Neural control

6.2 MODEL REFERENCE ADAPTIVE CONTROL (MRAC)

The basic block diagram of the adaptive control scheme is shown in Fig 6.2

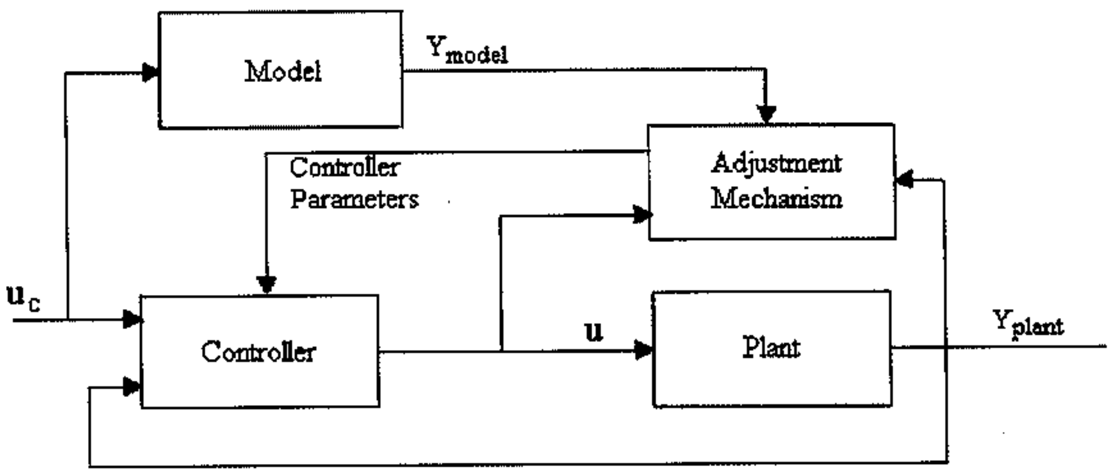


Figure 5.2 Basic block diagram of MRAC

The idea behind Model Reference Adaptive Control is to create a closed loop controller with parameters that can be updated to change the response of the system to match a desired model. MRAC begins by defining the tracking error, e. This is simply the difference between the plant output and the reference model output

$$e = y_{plant} - y_{model} \quad (6.1)$$

If a perfect model is assumed (plant = model), then the closed-loop system is stable if the controller C and the system are stable. However, under mismatch conditions (plant ‡ model), a low pass filter is introduced in the feed back loop to improve the controller robustness with respect to modeling errors.

6.3 SPEED CONTROLLER DESIGN

In Adaptive control method, The Model Referencing Adaptive Control Method is used. The model Referencing Adaptive control Scheme is shown in Fig 6.1.

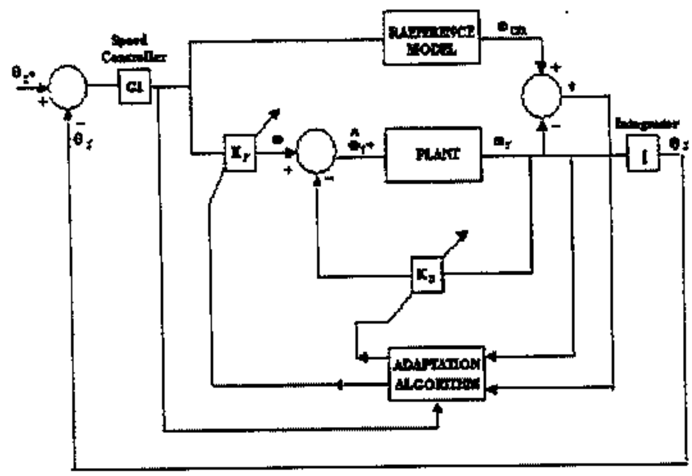


Figure 5.3. Model Referencing Adaptive speed control

In MRAC, as the name indicates, the plant's response is forced to track the response of the reference model, irrespective of the plant's parameter variation and the load disturbance effect. The reference model may be fixed or adaptive and is stored in any memory. The speed command ω_r^*, generated by the position control loop, is applied parallel to the reference model and plant controller. The reference model output speed ω_{rm} is compared with measured plant speed ω_r, and the resulting error *e* along with the ω_r signal actuates the adaptation algorithm. The feed forward and feed back gains K_F and K_B respectively, of the plant controller are iterated by the adaptation algorithm dynamically so as to reduce the error *e* to zero.

The adaptation algorithm can be defined as

$$\mathbf{K_F = K_{FO} + FV\omega_r^* + \int GV\omega_r^* \, dt} \quad (6.2)$$

$$\mathbf{K_B = K_{BO} + LV\omega_r + \int MV\omega_r \, dt} \quad (6.3)$$

$$\mathbf{V = D}e \quad (6.4)$$

Where K_{FO} and K_{BO} are the initial gain values and F, G, L, M and D are the adaptation law constants. The state equations can be given as

$$\mathbf{d/dt\,(\omega_{rm}) = A_m\omega_{rm} + B_m\,\omega_r^*} \quad (6.5)$$

$$\mathbf{d/dt\ (\omega_r) = A_P\omega_r + B_P\varpi_r} \tag{6.6}$$

And the other control loop equations are

$$\mathbf{\omega_r^* = G_1\ (\theta_r^* - \theta_r\ )} \tag{6.7}$$

$$\mathbf{\varpi_r^* = K_F\,\omega_r^* - K_B\omega_r} \tag{6.8}$$

$$e = \omega_{rm} - \omega_r \tag{6.9}$$

The control parameters K_F and K_B are time- varying. The stability of the system can be analyzed and correspondingly, parameters F, G, L, M and D can be determined.

7. SIMULATION RESULTS

7.1 SOFTWARE INTRODUCTION

MATLAB is a high-performance language for technical computing. It integrates computation, visualization, and programming in an easy-to-use environment where problems and solutions are expressed in familiar mathematical notation. Typical uses include Math and computation Algorithm development Data acquisition Modelling, simulation, and prototyping Data analysis, exploration, and visualization Scientific and engineering graphics Application development, including graphical user interface building MATLAB is an interactive system whose basic data element is an array that does not require dimensioning. This allows you to solve many technical computing problems, especially those with matrix and vector formulations, in a fraction of the time it would take to write a program in a scalar non-interactive language such as C or FORTRAN. The name MATLAB stands for matrix laboratory. MATLAB was originally written to provide easy access to matrix software developed by the LINPACK and EISPACK projects. Today, MATLAB engines incorporate the LAPACK and BLAS libraries, embedding the state of the art in software for matrix computation. MATLAB has evolved over a period of years with input from many users. In university environments, it is the standard instructional tool for introductory and advanced courses in mathematics, engineering, and science. In industry, MATLAB is the tool of choice for high-productivity research, development, and analysis. MATLAB features a family of add-on application-specific solutions called toolboxes. Very important to most users of MATLAB, toolboxes allow you to learn and apply specialized technology. Toolboxes are comprehensive collections of MATLAB functions (M-files) that extend the MATLAB environment to solve particular classes of problems. Areas in which toolboxes are available include signal processing, control systems, neural networks, fuzzy logic, wavelets, simulation, and many others.

7.2 FUZZY LOGIC TOOLBOX

The Fuzzy Logic Toolbox is a collection of functions built on the MATLAB numeric computing environment. It provides tools to create and edit fuzzy inference systems within the framework of MATLAB, or integrate your fuzzy systems into simulations with Simulink, or even build stand-alone C programs that call on fuzzy systems build with MATLAB. This toolbox relies heavily on Graphical User Interface (GUI) tools to accomplish work, although it can work entirely from the command line.

The toolbox provides three categories of tools:

- Command line functions
- Graphical, interactive tools
- Simulink blocks and examples

The first category of tools is made up of functions that can call from the command line or from applications. Many of these functions are MATLAB M-files, series of MATLAB statements that implement specialized fuzzy logic algorithms. MATLAB code for these functions can be viewed using the statement

type `function_name`

Secondly, the toolbox provides a number of interactive tools that access many of the functions through a GUI. Together, the GUI- based tools provide an environment for fuzzy inference system design, analysis, and implementation.

The third category of tools is a set of blocks for use with the Simulink simulation software. These are specifically designed for high speed fuzzy logic inference in the Simulink environment.

The vector control of the induction motor is developed using Simulink.

The SCIM parameters are as follows

Table 2. Machine Parameters

Pn	nominal power	3 HP
Rs	stator resistance	0.435Ω
Rr	rotor resistance	0.816Ω
Ls	stator inductance	0.002H
Lr	rotor inductance	0.002H
Lm	mutual inductance	69.31mH
J	rotor inertia	0.00819 kg.m2
2p	number of pole pairs	2

7.3. PI CONTROLLER

The parameters of the induction motor considered in this project are summarized in Table. 7.1. The performances of the controllers are evaluated separately under a variety of operating conditions.

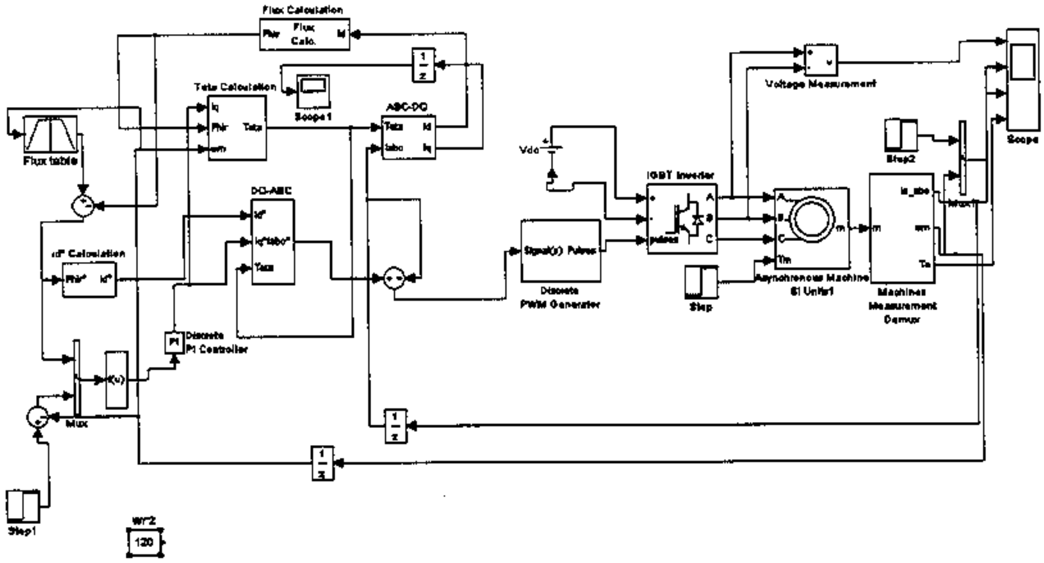


Fig. 7.1. Vector control of SCIM with PI controller

Fig. 7.1 shows the Field oriented control of induction motor with PI controller based speed controller design. In Fig.7.2 the speed response of the motor is shown for the step speed reference command. The figure shows the speed of the controller against the change in reference speed. The step speed command is given as a speed reference. The basic speed is 50 rad/s then the steady state speed is reached at 0.25 seconds and the starting torque is high. It is going up to 140 Nm. Then the speed is stepped to 100 rad/s at 0.3 seconds with in 0.15 seconds the steady state condition is reached. This shows the ability of the controller to reach the steady state than the conventional scalar controller. The conclusion from fig. 7.1 is the response of the controller is high and the oscillations are high at the rotor speed at the time of reaching the steady state condition.

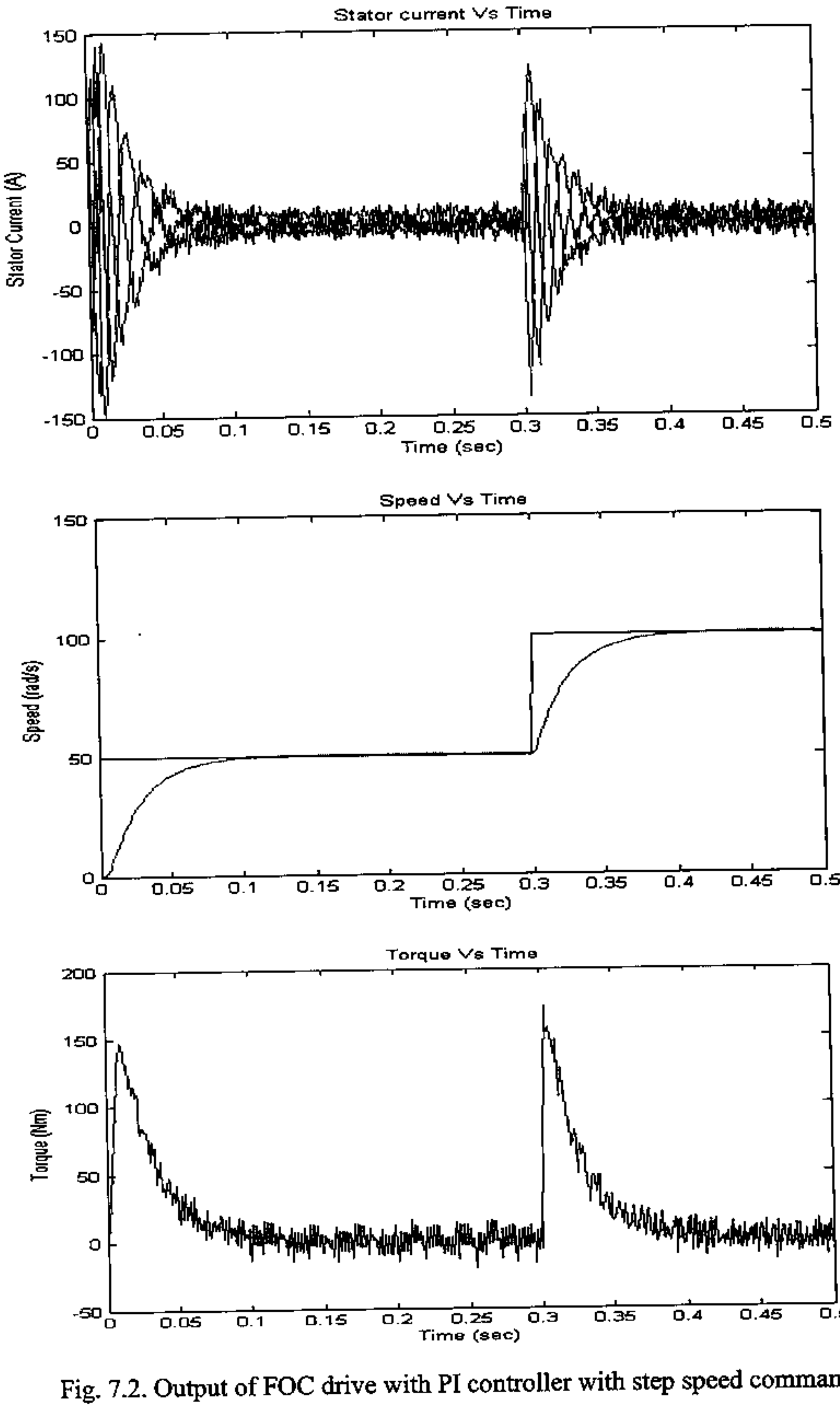


Fig. 7.2. Output of FOC drive with PI controller with step speed command

7.4. FUZZY LOGIC CONTROLLER

The main aim of the project is to improve the Robustness of the motor. So we go for the Fuzzy logic controller based vector control of induction motor. Fig 7.3 shows the fuzzy logic controller.

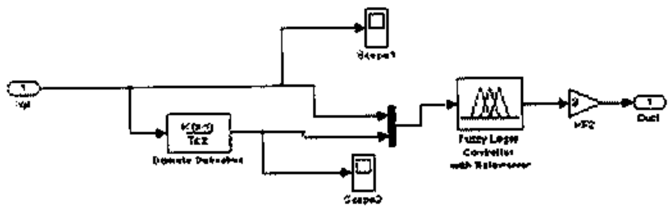


Fig. 7.3. Fuzzy logic controller

The performance of the fuzzy logic controller is shown in figure 7.4

Figure 7.4. Shows the performance of fuzzy logic controller at the load torque is 25 Nm as the initial load and step speed command is given as the speed reference. The performance of the controller is same as the PI controller but the starting torque is minimized by controlling the quadrature axis current Iq* and the starting torque is minimized to 50-70 Nm instead of 150 Nm at the Pi controller for the load Torque is below 10 Nm only. So the fuzzy logic controller is improving the robustness of the system. At the time of reaching the steady state condition, the torque is reaching its highest value as 70 Nm only. So the highest value of the torque is only 70 Nm for the whole time of operation.

Figure 7.5. Shows the performance of the controller when the load torque is 50 Nm. The step speed command is given as the speed reference. The response is same as the controller at 25 Nm load but the starting torque is slightly increased to 80 Nm from 40 Nm and the rotor is rotating at the negative speed for 0.05 seconds and the speed is increased to positive value. The same problem occurs at the time of reaching the steady state condition that the torque is reached to 120 Nm as the highest value. The speed reference is stepped to 120 Nm at the time of 2.5 seconds then the torque is increased to 80 Nm same as the initial torque reached at the time of starting. Finally the conclusion is by using the fuzzy logic controller the robustness of the system is improved.

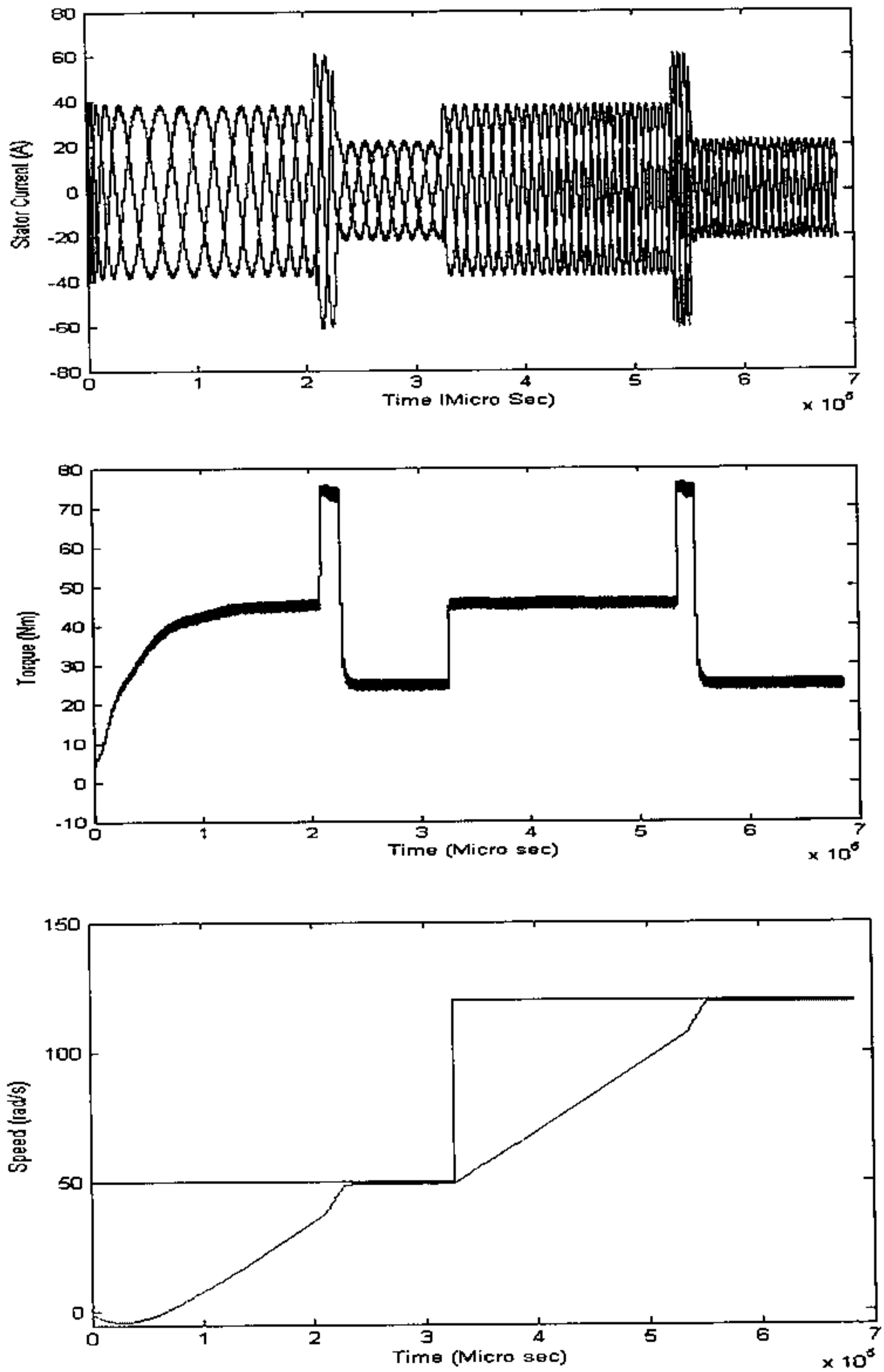


Fig. 7.4. Output of a FOC Drive with Fuzzy Logic Based Speed Controller at 25 Nm load

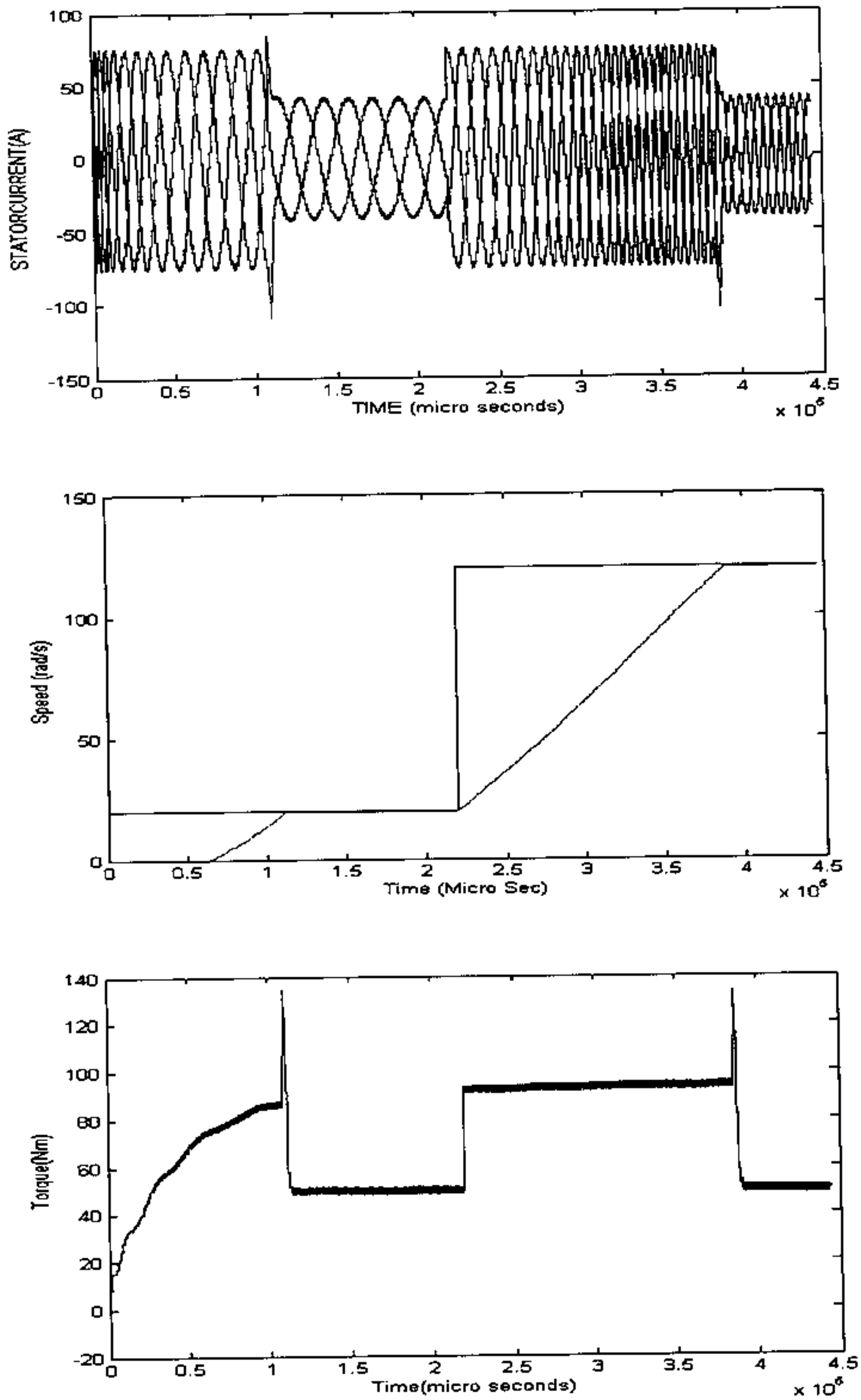


Fig 7.5. a) Stator Current in Amp b) Torque in Nm c) Speed in rad/s of the FOC Based Induction motor with Fuzzy Logic Based Speed Controller at 50 Nm load

7.5. MODEL REFERENCE ADAPTIVE CONTROLLER

When the load torque is applied to the motor accidentally, the MRAC is used to improve the response of the system. Fig 7.6 shows the model reference adaptive type speed controller.

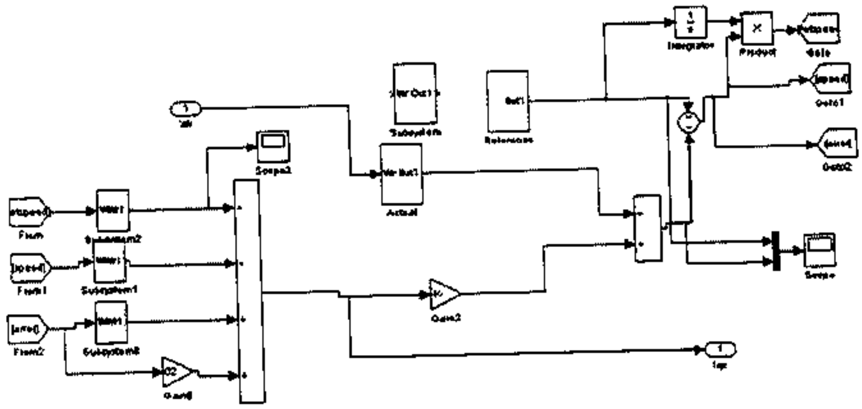


Fig 7.6. Model reference adaptive controller

Figure 7.6. Shows the response of the FOC drive with model reference adaptive controller based speed controller. The step speed command is given as the reference speed and the step load torque is given to the motor to check the response of the system with the immediate load torque. The response of the system is same as the fuzzy logic controller when the step speed command is given as the reference speed. When the load torque is stepped to 20 Nm at 0.5 seconds, the response of the system is very good and the speed is not affected by the variation of load and the torque is stepped down to low at 0.65 seconds and the speed is not affected. At last, the torque is stepped to 50 Nm at 0.7 seconds the speed is slightly down from the reference speed and the speed is increased to the reference speed at 0.1 seconds. From fig 7.6. The response of the MRAC system is calculated. The conclusion from the Figure is, for the step speed command, the response of the given system is same as the normal PI controller system. For the step load torque command the response of the given system is good compared to the conventional PI controller and the fuzzy logic controller. The speed reaches the steady state with in 0.1 seconds for the load torque is below 50 Nm.

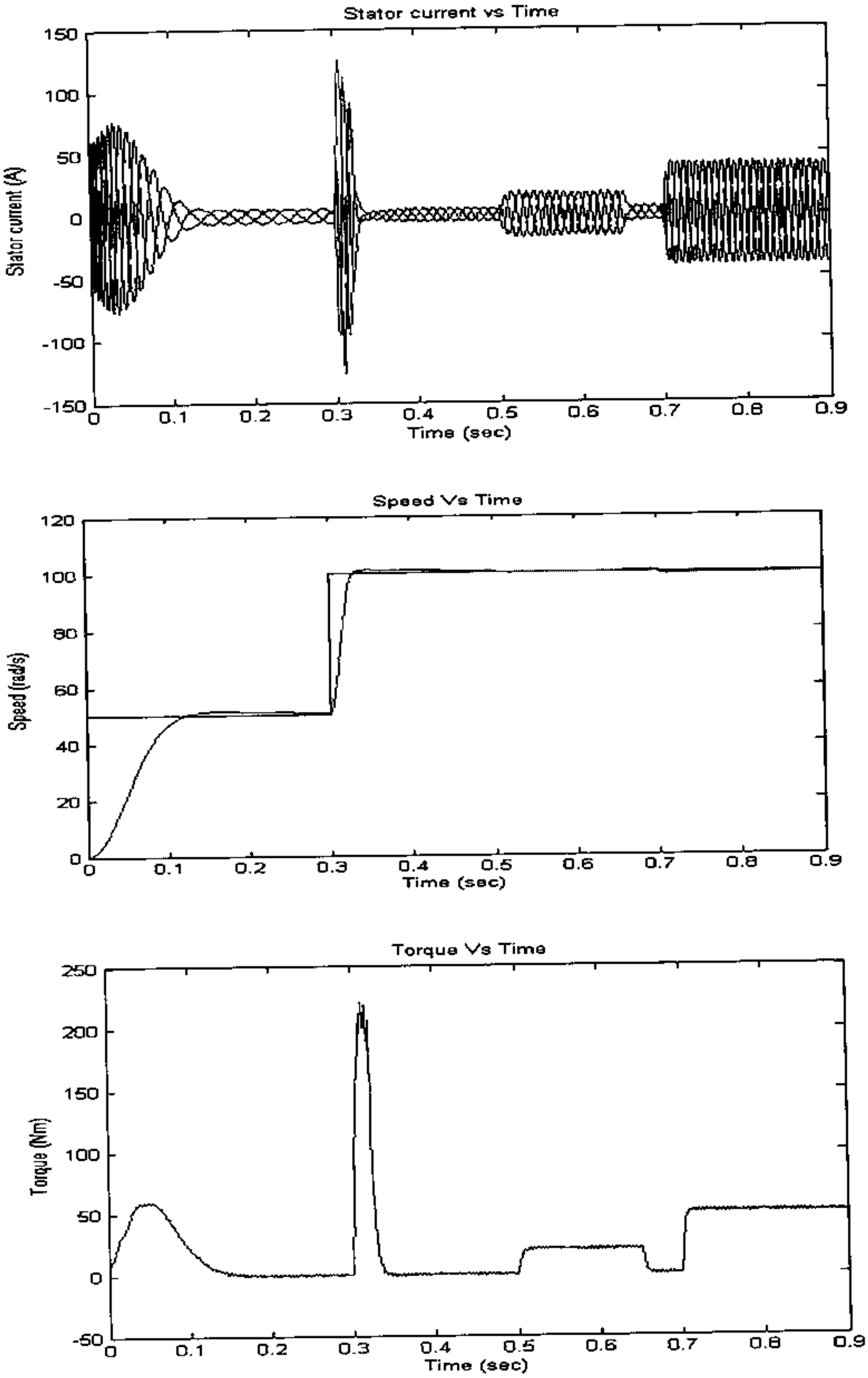


Figure 7.7. Output of FOC drive with model reference adaptive controller

8. RECOMMENDATIONS FOR FUTURE WORK

The Neural Networks can be used to optimize the output of Fuzzy Logic Controller to improve the Response of the system.

Fuzzy logic based Adaptive control can be used to improve the Response and Robustness of the existing Adaptive controller.

Various Types of switching Techniques like Space Vector Modulation can be used to minimize the switching losses in the Inverter.

9. CONCLUSION

The primary aim and objective of our project is to control the speed of Induction motor drives. Vector control technique is used to improve the dynamic performance of induction motor drives. By using the Indirect Vector Control Method, Steady state response of the given system could be achieved nearly at 0.25 seconds, which is quicker than scalar control techniques.

In this method the percentage error of speed is 0.1%, which is comparable to DC motor drives and by incorporating fuzzy logic approach, the starting torque of the motor, is minimised by controlling the quadrature axis current. By using adaptive controller the response of the controller of the system was improved. Steady state response of the given system is achieved nearly at 0.2 seconds, which is quicker than the conventional PI controllers and the speed reached the steady state condition with in 0.1 seconds when the load torque is stepped up 50 Nm which is faster than other controllers.

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