DIRECT TORQUE CONTROL OF INDUCTION MOTOR

A Project Report

Submitted by

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The above project has done by him in the Maintenance and Service Department of BHEL, Trichy -14, from 10-01-2006 to 10-04-2006, and found successful in his project work.

During this period he has shown keen interest in bearing various aspects of his project. The out come of this project is benefitable to our organization.

I wish them all success in future endeavors.

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ABSTRACT

In this project, a control scheme for torque and flux of induction motor based on stator flux estimation is discussed. The direct torque control (DTC) of an induction motor fed by a voltage source inverter is a simple scheme that does not need long computation time, can be implemented without mechanical speed sensors and is insensitive to parameter variation. In this principle, the motor terminal voltage and current are sampled and used to estimate the motor flux and torque. Based on the estimates of flux position and instantaneous error in torque and stator flux magnitude, a voltage vector is selected to restrict the torque and the flux errors within their respective torque and flux hysteresis band.

Direct Torque Control scheme employs the Space Vector Pulse Width Modulation (SVPWM) controller with fastest response and adjustable ripple rather than the conventional PWM controllers. DTC is simpler than the vector control method and less dependent on the motor model, because the stator resistance is the only machine parameter that is used for estimating the stator flux. The implementation of this scheme requires flux linkage and torque computations, pulse generation of switching states for inverter through a feed back control of the torque and flux directly without inner current loops. This results in reduced complexity of the system. The simulation results using MATLAB shows that the proposed scheme is not only effective in controlling the torque but also gives a better dynamic response and high efficiency than the conventional scalar control method.

ஆய்வு சுருக்கம்

இந்த ஆய்வானது நேரடி முறுக்கு விசை மற்றும் அதை சார்ந்த இயக்கு விசை கட்டுப்பாடு பற்றியதாகும். ஸ்டேடரின் இயக்கு விசையை அடிப்படையாக கொண்டு இண்டக்ஷன் மோட்டரின் கட்டுப்பாடு செயல்படுத்தபடுகிறது.

நேரடி முறுக்கு விசை கட்டுப்பாடு முறையானது இண்டக்ஷன் மோட்டரில் செயல்படுத்தப்படும் எளிமையான கட்டுப்பாட்டு முறையாகும் வேகத்தைக் கண்காணிக்கும் கருவிகள் இல்லாமை, குறைந்த கணக்கீட்டு நேரம் போன்றவை இந்த கட்டுப்பாட்டு முறையின் நிறைகள். இதில் இயந்திரத்தின் மின்னழுத்தம் மற்றும் மின்னோட்டம் ஆகியவற்றை கொண்டு இயந்திரத்தின் முறுக்கு விசை மற்றும் அதை சார்ந்த இயக்கு விசையை கொண்டு கணக்கிடப் படுகின்றது. இந்த கணக்கீடுகளின் அடிப்படையில் மின்னழுத்த கோள்கள் தேர்வு செய்யப்பட்டு அதன் மூலமாக முறுக்கு விசை மற்றும் இயக்கு விசை பிழைகள் குறைக்கப்படுகின்றன.

இந்த அமைப்பில் ஸ்பேஸ் வெக்டர் பல்ஸ் வித் மாடுலேஷன் (ப.வி.மா) கண்ட்ரோலர் உபயோகப்படுத்தப் படுகின்றது. அது பழைமையான பல்ஸ் வித் மாடுலேஷன் கண்ட்ரோலர்களைவிட சிறப்பானதாகும். இதில் ஸ்டேடர் இயக்கு விசை ஆனது, ஸ்டேடர் ரெசிஸ்டன்சை மட்டுமே கொண்டு கணக்கிடப்படுகின்றது. இதன் மூலம் மொத்த அமைப்பும் எளிமையாக்கப்பட்டுள்ளது. சிமுலேஷன் வெளியீடுகளின் மூலமாக இந்த முறையின் செயல்பாடு உறுதி செய்யப்படுகின்றது. இது ஸ்கேலார் கட்டுபாட்டு முறையைவிடவும் சிறப்பானது.

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LIST OF SYMBOLS AND ABBREVIATIONS

a	Complex operator $e^{3/2\pi j}$.
DTC	Direct torque control.
i_a, i_b, i_c.	Instantaneous values of stator phase a, b and c currents (A).
i_{qr},i_{dr}.	Instantaneous values of the rotor current dq-axes components in the stationary reference frame (A).
i_{qr}^r, i_{dr}^r rotor	Instantaneous values of the rotor current dq-axes components in the flux reference frame (A).
i_{qdr}.	Instantaneous values of the rotor current space vector in the stationary reference frame (A).
i_{qds}	Instantaneous values of the stator current space vector in the stationary reference frame (A).
J	Polar moment of inertia ($Kg{\cdot}m^2$).
L_M.	Mutual inductance per phase (H).
L_r.	Rotor self inductance per phase (H).
L_S	Stator self inductance per phase (H).
L_{lr}.	Rotor leakage inductance per phase (H).
L_{ls}.	Stator leakage inductance per phase (H).
MIPS	Million instructions per second.
P	Number of poles.
R_r.	Rotor resistance (Ohm).
R_s.	Stator resistance (Ohm).
T_e	Electromagnetic torque (N.m).
T_{ref}	Reference torque (N.m).

V_{qr}, V_{dr}.	Instantaneous values of the rotor voltage dq-axes components in the stationary reference frame (V).
V_{qs}, V_{ds}.	Instantaneous values of the stator voltage dq-axes components in the stationary reference frame (V).
V_{qdr}.	Instantaneous values of the rotor voltage space vector in the stationary reference frame (V).
V_{qds}.	Instantaneous values of the stator voltage space vector in the stationary reference frame (V).
λ_{qm} ,λ_{dm} .	Instantaneous values of the air gap flux linkage dq-axes components in the stationary reference frame (Wb).
λ_{qdm}.	Instantaneous values of the air gap flux linkage space vector in the stationary reference frame (Wb).
$\lambda_{qr}, \lambda_{dr}$.	Instantaneous values of the rotor flux linkage dq-axes components in the stationary reference frame (Wb).
$\lambda_{qs}, \lambda_{ds}$.	Instantaneous values of the stator flux linkage dq-axes components in the stationary reference frame (Wb).
λ_{qdr}.	Instantaneous values of the rotor flux linkage space vector in the stationary reference frame (Wb).
λ_{qds}	Instantaneous values of the stator flux linkage space vector in the stationary reference frame (Wb).
ω_r	Rotor speed (rad/sec).
T_{em}^*	Torque reference (Nm)
λ^*	Flux reference

CHAPTER I
INTRODUCTION

1.1. BACK GROUND

Before the introduction of microcontrollers and high switching frequency semiconductor devices, variable speed actuators were dominated by DC motors. Today, using modern high switching frequency power converters controlled by microcontrollers, the frequency, phase and magnitude of the input to an AC motor can be changed and hence the motor speed and torque can be controlled. AC motors combined with their drives have replaced DC motors in industrial applications due to their lower cost, better reliability, and lower weight and reduced maintenance requirement. Squirrel cage induction motors are more widely used than all the rest of the electric motors put together as they have all the advantages of AC motors and are easy to build.

1.2. BASIC AC MOTOR DRIVE SYSTEM

Fig.1.1 shows the block diagram of an AC motor drive system. A single phase or three phase AC power supply and an AC/DC converter provide a DC input to an inverter. A microcontroller decides the switching states for the inverter to control the motor's torque or speed. A sensing unit feeds back the terminal values such as motor speed, voltage and current to the microcontroller as needed for the closed loop control of the motor. Controllers used in AC motor drives are generally referred to as vector or field oriented controllers. The field oriented control methods are complex and are sensitive to inaccuracy in the motor parameters values.

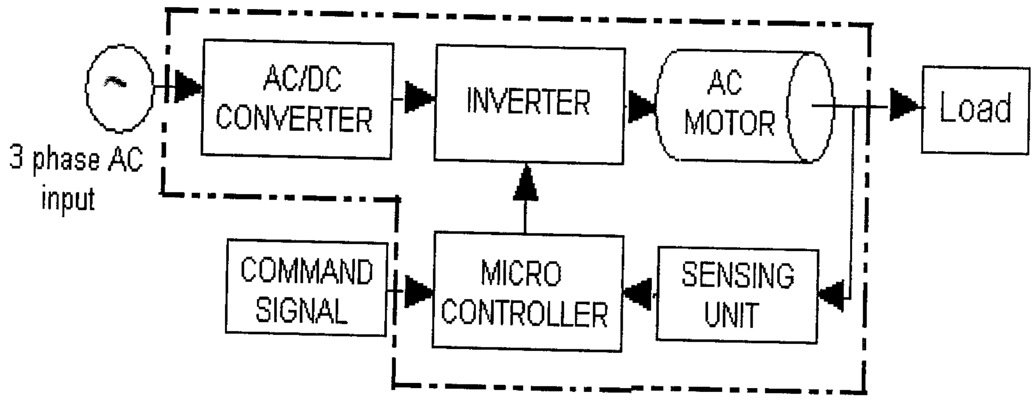


Fig.1.1. Block diagram of an AC motor drive system.

1.3. NEED FOR THE PROJECT

The high-speed induction motors like those that are implemented in electric drives possess high power to weight ratio. Such induction motors typically have low inductance and therefore needs a controller with a fast current response.

A suitable torque controller is required for these drives. Torque control is preferred for these applications instead of precise closed loop speed control because it minimizes the operation of non-linearities. It is important to make an electric drive like a standard drive.

To achieve this, a simplified variation of the field orientation known as direct torque control (DTC) was developed by Takahashi and Depenbrock. In direct torque controlled induction motor drives, it is possible to control directly the stator flux linkage and the electromagnetic torque by the selection of the optimum inverter switching state. The selection of the switching state is made to restrict the flux and the torque errors within their respective hysteresis bands and to obtain the fastest torque response and highest efficiency at every instant. DTC is simpler than field oriented control and is less dependent on the motor model, since the stator resistance value is the only parameter used to estimate the stator flux.

1.4 OBJECTIVE

The aim of my project is to reduce the torque ripple 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 space vector modulation (SVPWM) and hysteresis controller is used in this project.

1.5 ORGANIZATION OF THESIS

CHAPTER-1:

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

CHAPTER-2:

This chapter contains information about the constructional details, applications, advantages, disadvantages and control methods 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 field oriented control method.

CHAPTER-4:

This chapter contains information about the Direct Torque Control Method along with Space Vector Pulse Width Modulation Method.

CHAPTER-5:

This chapter deals about the Hardware description of this project and continue with simulation results.

CHAPTER II

CONTROL METHODS OF INDUCTION MOTOR

2.1 BASIC PRINCIPLES OF INDUCTION MOTOR

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 slots that the windings are positioned. The arrangement of the windings or coils within the stator determines the number of poles of the squirrel cage inductionmotor.

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 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 insinuated 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 fig. 2.1.

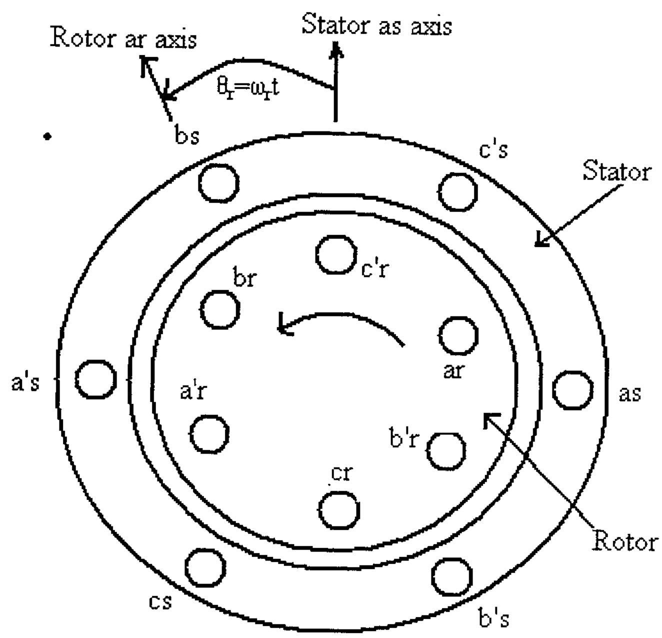


Fig 2.1 Symbolic representation of induction motor

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

A simple per phase equivalent circuit model of an induction motor is a very important tool for analysis and performance prediction at steady state condition. The synchronously rotating air gap flux wave generates a counter emf V_m, which is then converted to slip voltage $V_r' = nSV_m$ in rotor phase, where n = rotor to stator turns ratio and S = per unit slip. The stator terminal voltage V_s differs from voltage V_m by the drops in stator resistance R_s and stator leakage inductance L_{ls}. The excitation current I_o consists of two components: a core loss components $I_c = V_m/R_m$ and a magnetizing component $I_m = V_m/\omega_e L_m$, where R_m = equivalent resistance for core loss and L_m = magnetizing inductance. The rotor-induced voltage V_r' causes rotor current I_r' at slip frequency ω_{sl}, which is limited by the rotor resistance R_r' and the leakage reactance $\omega_{sl} L_{lr}'$. The stator current I_s consists of excitation components I_0 and the rotor-reflected current I_r, fig 2.2 shows the equivalent circuit with respect to the stator, where I_r is given as

$$I_r = nI_r' = (n^2 SV_m) / (R_r' + j\omega_{sl} L_{lr}') \qquad (1.1)$$

$$I_r = V_m / ((R_r/S) + j\omega_e L_{lr}) \qquad (1.2)$$

and parameters R_r ($=R_r'/ n^2$) and L_{lr} ($=L_{lr}' / n^2$) are referred to the stator. At standstill, S = 1, and therefore, fig 2.4 b corresponds to the short circuited transformer equivalent circuit. At synchronous speed, S = 0, current $I_r = 0$ and the machine takes excit6ation current I_0 only. At any sub synchronous speed, $0<S<1.0$, and with small value of S, the rotor current I_r is principally influenced by the R_r/S ($R_r/S >> j\omega_e L_{lr}$) parameter.

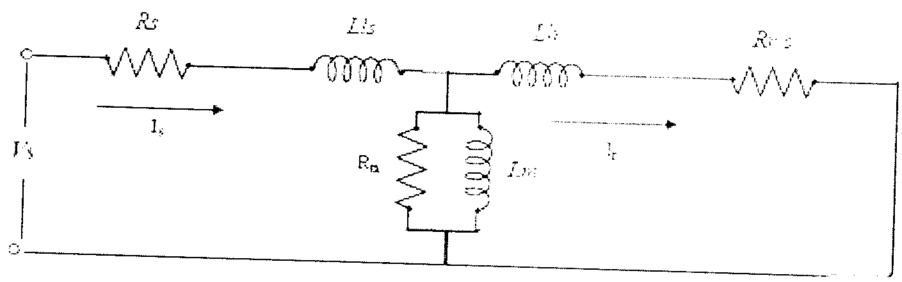


Fig.2.2 Per phase Equivalent circuit of induction motor

The phasor diagram for the equivalent circuit in fig 2.2 is shown in fig 2.3 where all the variables are in rms.

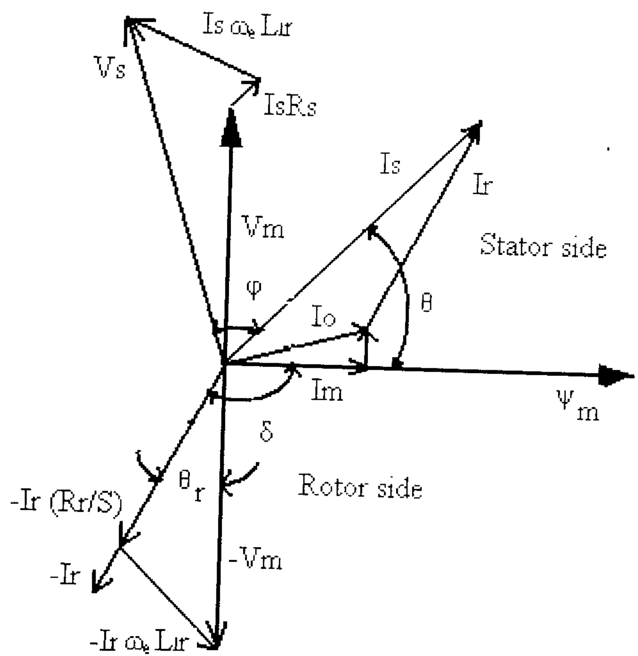


Fig 2.3 Phasor diagram for equivalent circuit of fig 2.2

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.

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
- Sensor less 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 CONTROL METHODS

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

1. Scalar Control

   a. Voltage fed inverter control

   - Open Loop Volts/Hz control
   - Energy conservation effect by variable frequency drive
   - Traction drive with parallel machines
   - Speed control with slip regulation
   - Speed control with torque and flux control
   - Current control voltage fed inverter drive

   b. Current fed inverter control

   - Independent Current & frequency control
   - Speed & Flux control in current fed inverter drive
   - Volts/Hz control of current fed inverter drive.

2. Vector Control

   a. Field oriented control

   - Direct field oriented control
   - Indirect field oriented control

   b. Direct Torque Control

2.6.1 Scalar control

The scalar control is based only on the magnitude variation of the control variables and it disregards the coupling effect in the machine. For example, the voltage of a machine can be controlled to control the flux, and frequency or slip can be controlled to control the torque. However, flux and torque are also functions of frequency and voltage respectively. Thus, it does not act on space vector position during transients.

The advantage includes,

i. It is simple and easy to implement.

ii. It finds application in all practical implementations.

The disadvantages includes,

i. It gives a sluggish response.
ii. The system is prone to instability because of a high order system effect.
iii. The temporary dip in flux reduces the torque sensitivity with slip and lengthens the response time.

2.6.2 Vector control

The vector control method is also known as decoupling, orthogonal or Tran vector control because it decouples the stator and rotor fluxes thereby giving DC machine like performance. This method is based on the relations valid in the dynamic states, not only in magnitude and frequency, but also on the instantaneous positions of voltage, current and flux space vectors. It acts on the positions of the space vectors and provides a better orientation in both steady state and during transients.

In vector control, the motor equations are transformed into field Co-ordinates that rotate in synchronism with the rotor flux vector. In the field Co-ordinates, under constant flux amplitude there is a linear relationship between the control variables and torque. Moreover, like in a separately excited DC motor, the reference for the flux amplitude is reduced in the field weakening region in order to limit the stator voltage at high speed.

CHAPTER III
VECTOR CONTROL OF INDUCTION MOTOR

3.1 DQ MODEL OF INDUCTION MACHINE

The vectors are represented in the stationary dq reference frame. By using instantaneous vectors, the behavior of an induction machine can be conveniently

$$\begin{bmatrix} R_s + L_s p & 0 & L_m p & 0 \\ 0 & r_s + L_s p & 0 & L_m p \\ L_m p & \omega_r L_m & r_s + L_r p & \omega_r L_r \\ -\omega_r L_m & L_m p & -\omega_r L_m & r_s + L_r p \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{qr} \\ i_{dr} \end{bmatrix} = \begin{bmatrix} V_{qs} \\ V_{ds} \\ V_{qr} \\ V_{dr} \end{bmatrix}$$

(2.1 a)

The dq components of a balanced three phase voltages and currents are calculated by substitution in Equation 2.1b.

$$\begin{bmatrix} f_{qs} \\ f_{ds} \end{bmatrix} = \begin{bmatrix} 2/3 & -1/3 & -1/3 \\ 1/\sqrt{3} & -1/\sqrt{3} & 0 \end{bmatrix} \begin{bmatrix} f_a \\ f_b \\ f_c \end{bmatrix}$$

(2.1 b)

Equation 2.1a can be expressed in complex notation using the definitions

$$i_{qds} = i_{qs} - j\, i_{ds} \qquad i_{qdr} = i_{qr} - j i_{dr}$$

$$V_{qds} = V_{qs} - jV_{ds} \qquad V_{qdr} = V_{qr} - V_{dr} = 0$$

(2.2)

Equation 2.3 is the equivalent complex form of Equation 2.1a.

$$\begin{bmatrix} V_{qds} \\ V_{qdr} \end{bmatrix} = \begin{bmatrix} r_s + L_s p & L_m p \\ L_m (p - j\omega_r) & r_r + L_r (p - j\omega_r) \end{bmatrix} \begin{bmatrix} i_{qds} \\ i_{qdr} \end{bmatrix}$$

(2.3)

In a symmetrical, three phase induction machine the instantaneous electromagnetic torque can be written as:

$$T_e = (3/2)\,(P/2)\,\mathrm{Im}\,\lambda_{qds}^{*}\, iqdr$$

$$T_e = (3/2)\,(P/2)\,(L_m/L_r)\,\mathrm{Im}\,\lambda_{qdr}^{*}\, i_{qds}$$

(2.4)

3.2 FIELD ORIENTED VECTOR CONTROL OF INDUCTION MOTOR

Field oriented control refers to induction motor operation in a synchronously rotating dq reference frame that is aligned with one of the motor fluxes, typically the rotor flux .in this mode of operation, control of the torque and flux is decoupled such that the d-axis component of the stator current controls the rotor flux magnitude and the q-axis component controls the output torque. The required d- axis component of the stator current, ids, to achieve a given rotor flux magnitude demand, λdr, can be determined from equation

$$\lambda_{dr} = L_m i_{ds} \qquad (2.5)$$

The required q-axis component of the stator current, iqs for a given torque demand Tem*, can be determined from

$$T_{em}^{*} = (3/2)(P/2)(L_m/L_r)\lambda_{dr}^{*} i_{qs} \qquad (2.6)$$

In this control a 90- degree orientation between the rotor flux and the q-axis component of the stator current is maintained thus the machine operate in orientation similar to the DC machine and the torque will follow the magnitude of the q-axis current component . If a constant rotor flux magnitude is maintained aligned with the d- axis Equation 2.4 modifies to

$$Te = (3/2)(P/2)(L_m/L_r)\lambda_r i_{qs} \qquad (2.7)$$

For the d axis of the synchronously rotating reference frame to be aligned with the rotor flux the 'slip relation', must be maintained as

$$\omega_e - \omega_r = (L_m/L_r)(i_{qs}/i_{ds}) \qquad (2.8)$$

FOC consists of a proportional- integral controller that regulates the stator voltage to achieve the calculated stator current. The required voltage is then synthesized by the inverter using SVM.

During the motor operation the actual rotor resistance and inductance can vary, for example with temperature. The resulting errors between the values used and the actual parameters cause an incomplete decoupling between torque & flux.

To keep the 90-degree orientation the rotor flux position feedback is needed. There are two methods that are used to provide the feedback signal namely, direct field, and indirect field oriented control.

3.2.1 Direct field oriented control

In direct field-oriented control , the rotor flux position is obtained by measuring the air gap flux using sensors or estimated using terminal voltages and currents, which is also known as sensor less direct field oriented control. Knowing the flux position the required phase currents that give the desired flux and torque value can be calculated and forced into the stator of the induction motor.

Flux sensors provide the value of the air gap flux λ_{qdm}, from which the rotor flux position can be calculated using Equations 2.9a-c.

$$\lambda_{dr} = L_r/L_m\ \lambda_{dm} - L_{lr}\ i_{ds} \qquad (2.9\ a)$$

$$\lambda_{qr} = L_r/L_m\ \lambda_{qm} - L_{lr}\ i_{qs} \qquad (2.9\ b)$$

$$\theta_{rf}{}^* = L_r/L_m\ \lambda_{qm} - L_{lr}\ i_{qs} \qquad (2.9\ c)$$

Where,

$$L_{lr} = L_r - L_m,\ i_{qs} = i_a \text{ and } i_{ds} = -1/\sqrt{3}\ i_a - 2/\sqrt{3}\ i_b$$

Knowing θrf^* the current command is transformed from the rotor flux reference frame to the stationary reference frame using equation 2.10 a-b and the required phase current commands are calculated using equations 2.11 a-c.

$$i_{qs}{}^* = i_{qs}{}^{r*} \cos\theta_{rf}{}^* + i_{ds}{}^{r*} \sin\theta_{rf}{}^* \qquad (2.10\ a)$$

$$i_{ds}{}^* = -\ i_{qs}{}^{r*} \sin\theta_{rf}{}^* + i_{ds}{}^{r*} \cos\theta_{rf}{}^* \qquad (2.10\ b)$$

$$i_a{}^* = iqs^* \qquad (2.11\ a)$$

$$i_b{}^* = -1/2\ i_{qs}{}^* - \sqrt{3}/2\ i_{ds}{}^* \qquad (2.11\ b)$$

$$i_c{}^* = 1/2\ i_{qs}{}^* + \sqrt{3}/2\ i_{ds}{}^* \qquad (2.11\ c)$$

The flux sensors can be avoided by estimating the rotor flux position from the terminal voltage and currents. First, the stator flux is estimated using equations 2.12 a & 2.12 b

$$\lambda_{qs} = \int (Vqs - r_s\ i_{qs})\ dt \qquad (2.12\ a)$$

$$\lambda_{ds} = \int (Vds - r_s\ i_{ds})\ dt \qquad (2.12\ b)$$

then the rotor flux dq components are calculated using equations 2.13a- 2.3b

$$\lambda_{qr} = L_r/L_m\ (\lambda_{qs} - L_s'\ i_{qs}) \qquad (2.13\ a)$$

$$\lambda_{dr} = L_r/L_m\ (\lambda_{ds} - L_s'\ i_{ds}) \qquad (2.13\ b)$$

Where,

$L_s' = L_s - L_m^2/L_r$, the rotor flux position is calculated using equation 2.9 c. The Fig 2.2 shows the direct field oriented control method.

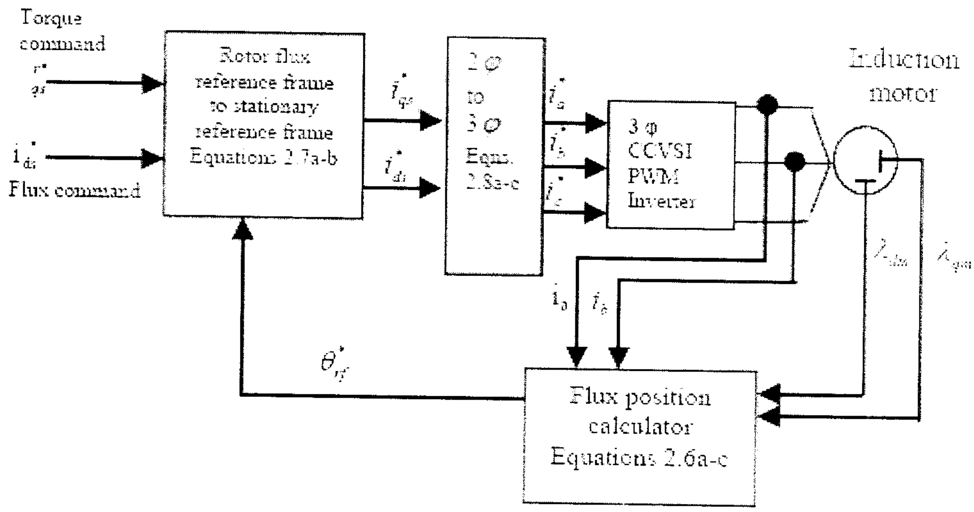


Fig.3.1 Block diagram of direct field oriented control methoc.

3.2.2 Indirect Field-Oriented Control

In indirect field-oriented control, the 90-degree orientation between the rotor flux and the q-axis component of the stator current is maintained by satisfying the unique slip relationship associated with the dq currents components in a field-orientation controlled motor (Equation 2.11).

$$\omega_e - \omega_r = S\omega_e^* = (r_r / L_r)(i_{qs}^{r*} / i_{ds}^{r*}) \qquad (2.14)$$

In this method, a feedback of the rotor position is required. Similar to the rotor flux position in direct field-oriented control case the rotor position can be directly obtained from an incremental encoder or estimated using the terminal voltages and currents information which is known as sensor less indirect field-oriented control. The slip frequency is estimated using Equation 2.14.The rotor flux position is obtained by summing the rotor position and the slip position as in Equation 2.15

$$\theta_{rf}^* = \int S\omega_e^* \, dt + \theta_m = \theta_s^* + \theta_{rm} \qquad (2.15)$$

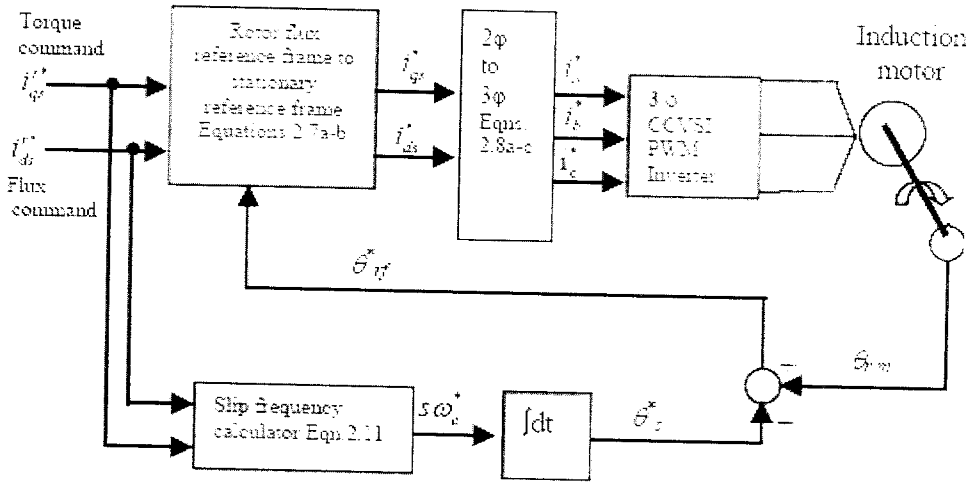


Fig.3.2 Block diagram of indirect FOC

3.3 LIMITATIONS OF FOC

Field oriented control in induction motor is sensitive to deviation in estimated motor parameter values used in the control algorithm from their actual values. In addition, the parameter values may need to be updated while the motor is running, if there are large changes in the operating temperature and inductance values changes due to flux saturation. The desired field orientation torque characteristic cannot be achieved unless the controller provides real time tuning for the parameter values.

Direct field oriented control requires,

i. The values of the rotor leakage inductance, (L_{lr}) and the ratio of the rotor inductance to the mutual inductance (L_r/L_m), while the first parameter has a a constant value independent of temperature and flux, the second parameter is moderately affected by the saturation of the main flux path in the motor.

ii. The special flux sensors, which need frequent maintenance and impose limitation in the motor mounting.

iii. In case of senseless form, the values of stator resistance and inductance are needed which are sensitive to temperature changes.

iv. If sensors are used to measure the flux, the parameters values mismatch will affect the steady state torque values but not the dynamic response and if flux is estimated then the dynamic response is deteriorated.

Fig 2.4 shows the effects of 10% over estimation of the values of the motor inductance (Llr + Lr/Lm) and 10 % underestimation of the values of the rotor resistance (rr) on the performance of the motor.

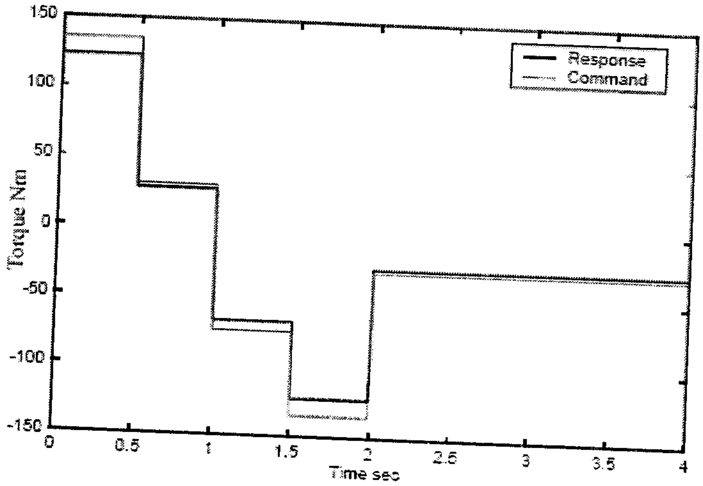


Fig.3.3 Sensitivity of direct oriented control to inaccuracy in parameter values

Indirect field orientation is more sensitive to parameters values inaccuracy than direct field orientation. The value of the rotor time constant (L_r/r_r) is used in the slip frequency calculation this parameter is sensitive to temperature and flux level. The need of special position incremental encoder is another disadvantage of the method. Fig. 2.4 shows the effect of 10% overestimation of the values of the motor inductances (L_{lr} and L_r/L_m) and 10% underestimation of the values of the rotor resistance (r_r) in the performance of an indirect field-oriented controller which is assumed to be able to force the exact torque and flux torque commands into the motor.

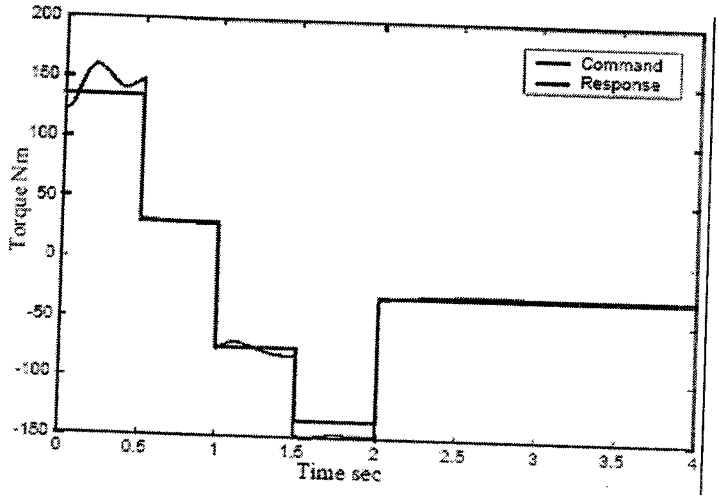


Fig.3.4 Sensitivity of indirect field oriented control to inaccuracy in parameters

Beside the above limitations, field-oriented control is complex and need fast speed micro-controllers. These limitations make an alternative method known, as direct torque control an attractive method because

i. It has a simple algorithm.
ii. It does not need phase transformation.
iii. It does not need special position or flux sensors, and
iv. It needs only resistance of the machine parameter.
v. It does not need any current controller because it select the voltage space vectors according to the error in flux linkage torque.

CHAPTER IV
DIRECT TORQUE CONTROL

4.1. CONCEPT OF DTC

In three phase induction motor drives a complete decoupling of flux and torque control variables is usually required. The torque command is generated from speed loop controller or directly by the user. The flux command is selected according to operation requirements, i.e., field weakening operation to achieve a wide speed range or flux regulation in accordance with load conditions to minimize motor losses.

In most control strategies the input commands are torque and flux, whereas the output commands are three phase reference currents. Then, the power converter has the ability to force only designed current waveform in to stator windings. For this purpose the current controlled VSI can be used. In order to obtain good performance the desired current waveform has to be well approximated leading to high switching frequency.

DTC uses an induction motor model to predict the voltage required to achieve a desire output torque. By using only current and voltage measurements, it is possible to estimate the instantaneous stator flux and output torque. An induction motor model is then used to predict the voltage required to drive the flux and torque to the demanded values with in a fixed time period.

Fig 4.1 shows the block diagram of DTC, the voltage vector required to drive the error in the torque and flux to zero is calculated directly. If the inverter is not capable of generating the required voltage, then the voltage vector which will drive the torque and flux towards the demand values is chosen and held for the complete cycle.

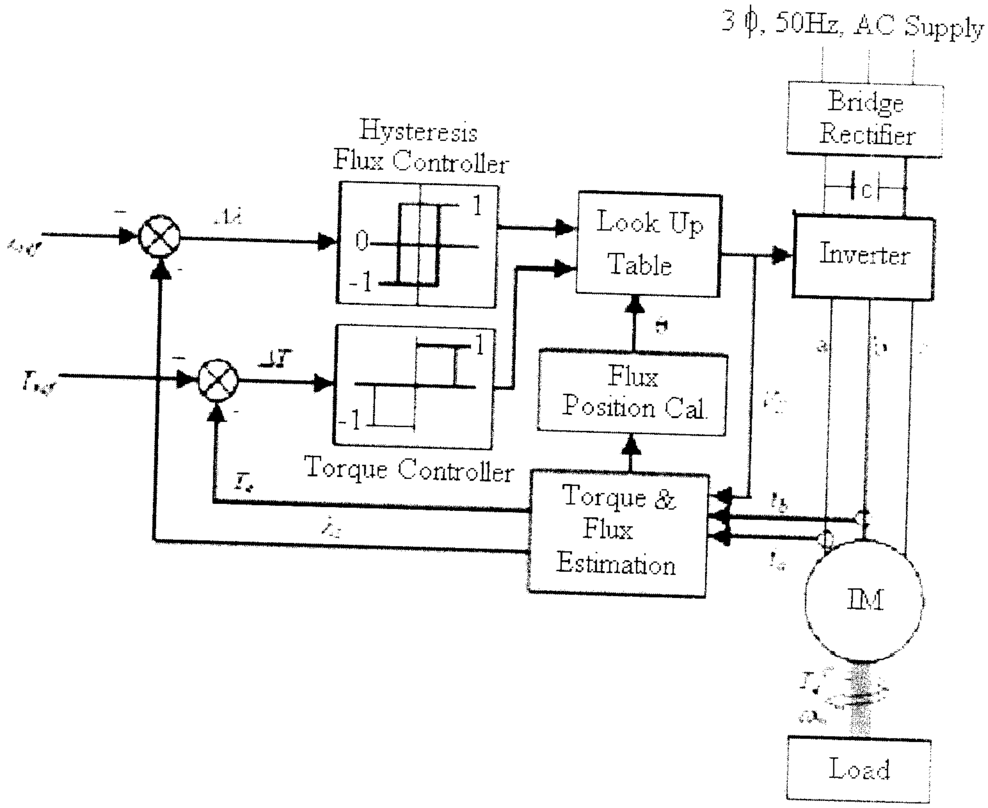


Fig 4.1 General Block Diagram of DTC.

In DTC induction motor drive supplied by a VSI, it is possible to control directly the stator flux linkage λ_s & the electromagnetic torque by the selection of an optimum inverter voltage vector. The selection of the voltage vector of the VSI is made to restrict the flux & torque error with in their respective flux & torque hysteresis band and to obtain fastest torque response and highest efficiency at every instant. It enables both quick torque response in the transient operation and reduction of the harmonic losses & acoustic noise.

4.2. TORQUE EXPRESSION WITH STATOR AND ROTOR FLUXES

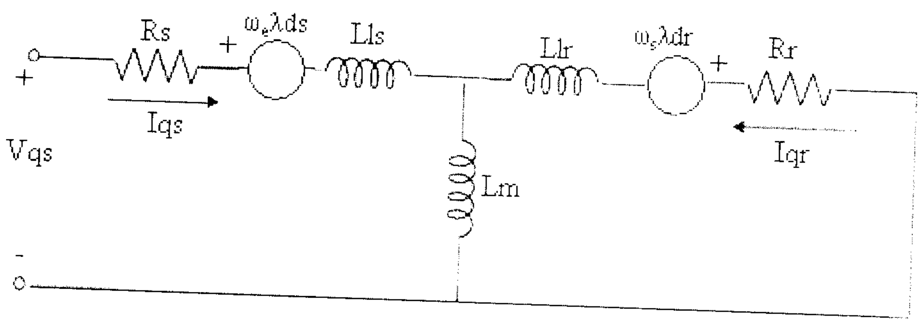
The torque expression given in equation $T_e = (3/2)\,(P/2)\,(\psi_{ds}^{s}\, i_{qs}^{s} - \psi_{qs}^{s}\, i_{ds}^{s})$ can be expressed in the vector form as

$$T_e = (3/2)\,(P/2)\,\psi_s\, I_s \qquad (4.1)$$

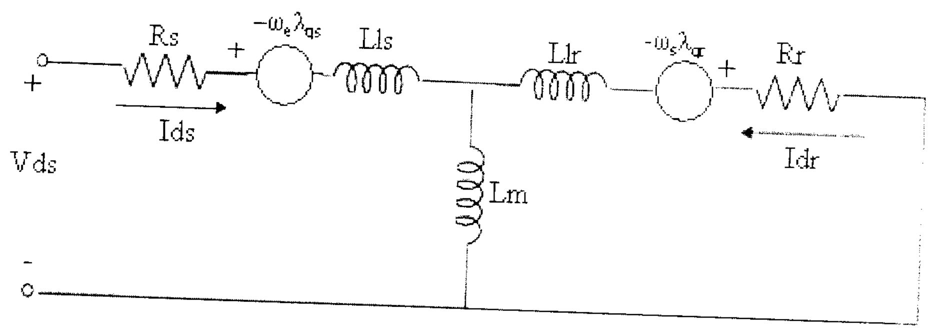
Where $\psi_s = \psi_{qs}^s - \psi_{ds}^s$ and $I_s = i_{qs}^s - j\, i_{ds}^s$. In this equation, I_s is to be replaced by rotor flux ψ_r. In the complex form, ψ_s and ψ_r can be expressed as function of current (from equivalent circuit) as

$$\psi_s = L_s I_s + L_m I_r \qquad (4.2)$$

$$\psi_r = L_r I_r + L_m I_s \qquad (4.3)$$



A. q – axis equivalent circuit



B. d – axis equivalent circuit

Fig.4.2 Equivalent circuit of an IM in a dq reference frame

Eliminating I_r, from equation 4.2, we get

$$\psi_s = (L_m / L_r)\, \psi_r + L_s' I_s \qquad (4.4)$$

Where $L_s' = L_s L_r - L_m^2$. The corresponding expression of Is is

$$I_s = (1/ L_s')\, \psi_s - (L_m / L_r L_s')\, \psi_r \qquad (4.5)$$

Substituting equation (4.5) in equation (4.1) and simplifying yields

$$T_e = (3/2)\,(P/2)\,(L_m / L_r L_s')\, \psi_r\, \psi_s \qquad (4.6)$$

That is, the magnitude of torque is

$$T_e = (3/2)\ (P/2)\ (L_m / L_r L_s')\ |\psi_r|\ |\psi_s| \sin \gamma \qquad (4.7)$$

Where γ is the angle between the fluxes, fig.4.3 shows the phasor (or vector) diagram for equation (4.6), indicating the vector ψ_r, ψ_s, and I_s for positive developed torque. If the rotor flux remains constant and the stator flux is changed incrementally by stator voltagr V_s as shown, and the corresponding change of γ angle is Δr, the incremental torque ΔT_e expression is given as

$$\Delta T_e = (3/2)\ (P/2)\ (L_m / L_r L_s')\ |\psi_r|\ |\psi_s + \Delta\psi_s| \sin \Delta\gamma \qquad (4.8)$$

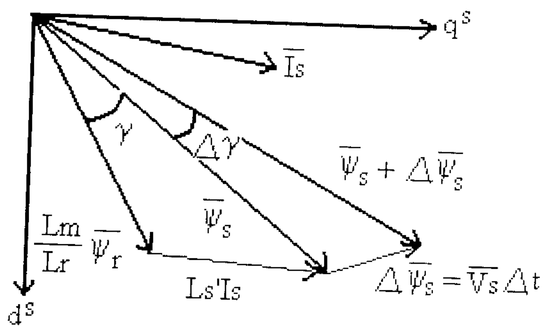


Fig 4.3 Stator flux, rotor flux, and stator current vector on $d^s - q^s$ plane

4.3. VOLTAGE SOURCE INVERTER

There are many topologies for the VSI used in DTC control of induction motor that gives high number of possible number of possible output voltage vectors but the most common one is the six step inverter. A six step voltage source inverter provides the variable frequency AC voltage input to the induction motor in the direct torque control method. The DC supply to the inverter is provided either by a DC source like a battery, or a rectifier supplied from a three phase AC source. Fig.4.4. shows the block diagram of a six step voltage source inverter. The inductor L is used to limit short through fault current. A large electrolytic capacitor C is inserted to stiffen the DC link voltage. The switching devices employed in the voltage source inverter bridge must be capable of instantaneous turn on and off. To achieve this, the IGBT's are used because they offer high switching speed with enough power rating. Each IGBT has an inverse parallel connected diode. This diode provides alternate path for the motor current after the IGBT is turned off.

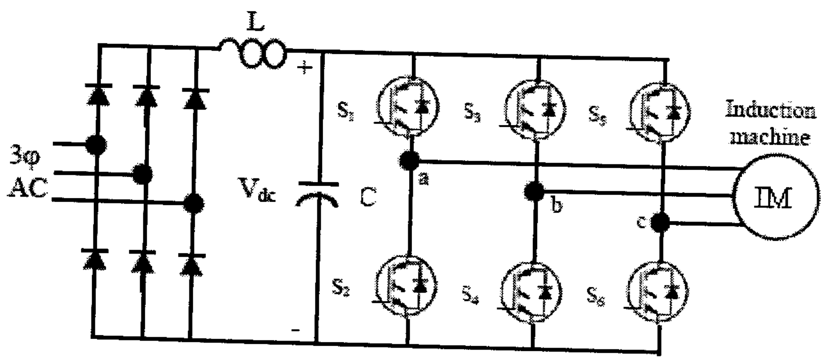


Fig.4.4 Block diagram of a six step voltage source inverter.

Each leg of the inverter has two switches, one connected to the positive terminal of the DC link and the other to the negative terminal. Only one of the switches in the leg is turned on at any instant.

Considering the combinations of the status of phases a, b, c, the inverter has eight switching states (VaVbVc=000-111), in which the two are zero vectors V_0 (000) and V_7 (111) where the motor terminals are short circuited and the other six voltage vectors are non-zero voltage vectors, from V1 to V6. Each vector lies in the center of a sector of 60-degree width named as S1 to S6 according to the voltage vector.

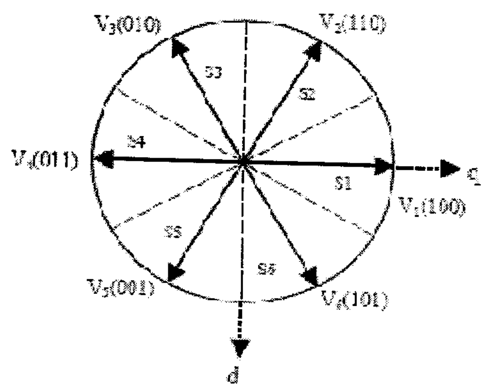


Fig. 4.5 Voltage vectors for the VSI switching states

4.4. PULSE WIDTH MODULATION

The inverter is simple and the switching loss is low because there are only six switching per cycle of fundamental frequency. Unfortunately, the lower harmonics of the six-step voltage wave will cause large distortions of the current wave unless filtered by bulky and uneconomical low-pass filters. Besides, the voltage control by the line-side rectifier has the usual disadvantages.

PWM principle

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.

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 space-vector PWM (SVM) method is an advanced, computation PWM method and is possibly the best among all the PWM techniques for variable frequency drive applications. Because of its superior performance characteristics, it has been finding widespread application in recent years.

4.4.1. Space Vector PWM (SVPWM)

The PWM method is discussed so for have only considered implementation on a half bridge of a three phase bridge inverter. If the load neutral is connected to the centre tap of the dc supply, all three-bridges operate independently, giving satisfactory PWM performance. With a machine load, the load neutral is normally isolated, which causes interaction among the phases. This interaction was not considered before in the PWm discussion. The SVM method considers this interaction of the phases and optimizes the harmonic content of the three phase isolated neutral load. For example, if the three-phase sinusoidal and balanced voltages given by the equation

$$V_a = V_m \cos\omega t \quad (4.9)$$

$$V_b = V_m \cos(\omega t - (2\pi/3)) \quad (4.10)$$

$$V_c = V_m \cos(\omega t + (2\pi/3)) \quad (4.11)$$

Are applied to a three phase induction motor, using equation $V=2/3[v_{as} + av_{bs} + a^2v_{cs}]$, it can be shown that the space factor With magnitude V_m rotates in a circular orbit at angular velocity ω where the direction rotation depends on the phase sequence of the voltages. With the sinusoidal three phase command voltages, the composite PWM fabrication at the inverter output should be such that the average voltage follows these command voltages with a minimum amount of harmonic distortion.

Converter Switching States

A three phase bridge inverter, as shown in figure. Have $2^3 = 8$ permissible switching states. Table 4.1 gives a summary of the switching states and corresponding phase to neutral voltage of an isolated neutral machine.

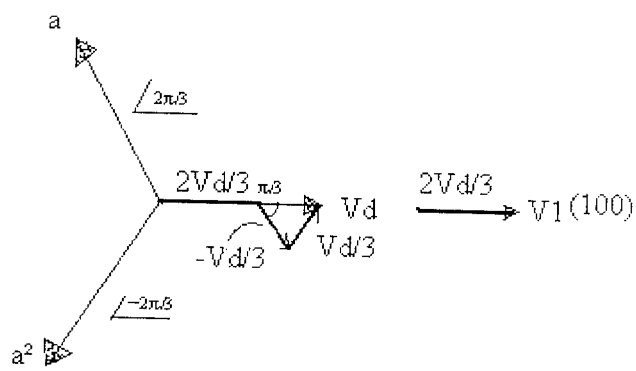


Fig 4.6 Construction of inverter space vector V_1 (100)

Table 1 Summary of inverter switching states

State	On devices	Van	Vbn	Vcn	Space voltage vector
0	$Q_4Q_6Q_2$	0	0	0	$V_0(000)$
1	$Q_1Q_6Q_2$	$2V_d/3$	$-V_d/3$	$-V_d/3$	$V_1(100)$
2	$Q_1Q_3Q_2$	$V_d/3$	$V_d/3$	$-2V_d/3$	$V_2(110)$
3	$Q_4Q_3Q_2$				$V_3(010)$
4	$Q_4Q_3Q_5$				$V_4(011)$
5	$Q_4Q_6Q_5$				$V_5(001)$
6	$Q_1Q_6Q_5$				$V_6(101)$
7	$Q_1Q_3Q_5$	0	0	0	$V_7(111)$

Consider, for example, state 1, when switches Q_1, Q_6, and Q_2 are closed. In this state, phase a is connected to the positive bus and phases b and c are connected to the negative bus. The simple circuit solution indicates that $v_{an} = 2/3v_d$, $v_{bn} = -1/3v_d$, and $v_{cn} = -1/3v_d$. The inverter has six active states (1-6) when the voltage is impressed across the load, and two aero states (0 and 7) when the machine terminals are shorted the lower devices or upper devices, respectively. The sets of phase voltages for each switching state can be combined with the help of equation $V = 2/3[v_{as} + av_{bs} + a^2v_{cs}]$ to derive the corresponding space vectors. The graphical derivation of V_1 (100) in fig 4.7 indicates that the vector has a magnitude of $2/3V_d$ and is aligned in the horizontal direction as shown.

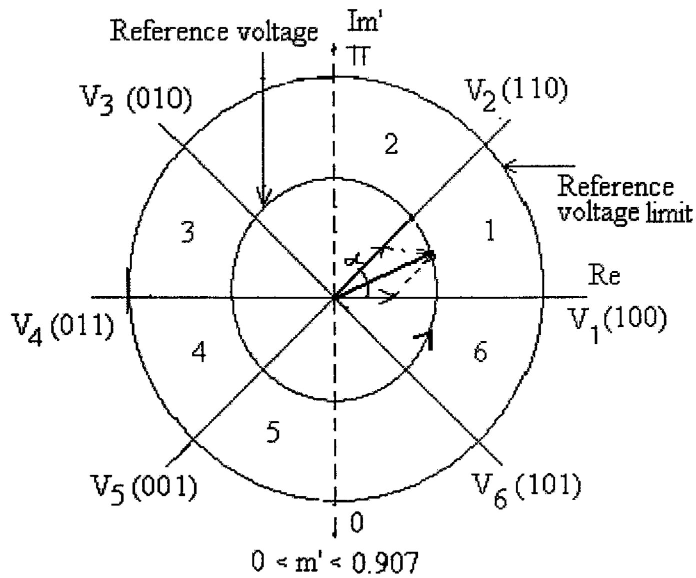


Fig 4.7 Space vectors of three-phase bridge inverter showing reference voltage trajectory and segment of adjacent voltage vectors

In the same way, all six active vectors and two zero vectors are derived and plotted in fig 4.7. The active vectors are p/3 angle apart and describe a hexagon boundary (shown as dotted). The two zero vectors V_0 (000) and V_7 (111) are at the origin. For three-phase, square wave operation of the inverter. As shown in fig 4.7, it can be easily verified that the vector sequence is V_1? V_2? V_3? V_4? V_5? V_6, with each dwelling for an angle of p/3, and there are no zero vectors.

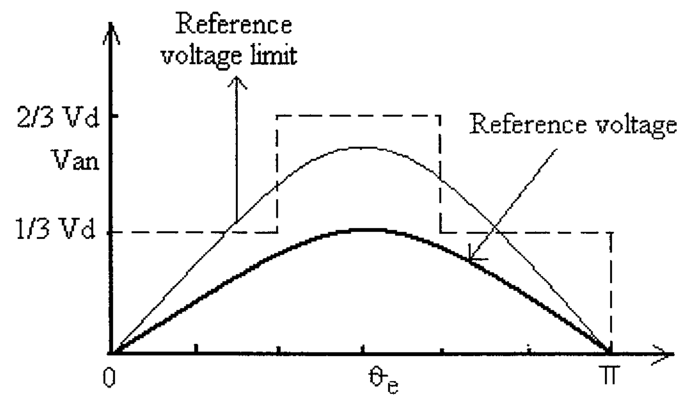


Fig 4.8 Corresponding reference phase voltage wave

4.5 STATOR FLUX BASED DTC

Since the stator flux based direct torque control will be used in the thesis a detailed description of the method is presented. In direct torque control fast torque response can be obtained by selecting the optimal VSI switching state if the flux magnitude is kept constant can be expressed as

$$T_e = (3/2)\,(P/2)\,|\lambda_{qds}|\,|i_{qds}|\,\sin(\sigma_s - k_s) \qquad (4.12)$$

Where, $\sigma_s - k_s$ is the angle between the stator flux linkage and stator current space vector as shown in fig.4.9

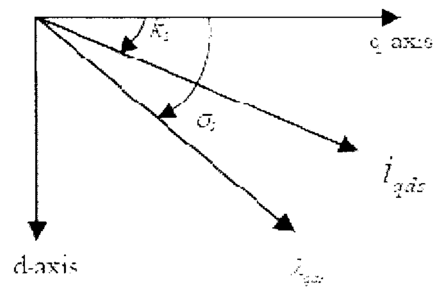


Fig 4.9 stator flux linkage and stator current space vectors.

The rotor variables in equation (4.13) can be eliminated to get equation (4.14)

$$\lambda_{qds} = L_s\, i_{qds} + L_m\, i_{qdr}$$

$$\lambda_{qdr} = L_r/L_m\,(\lambda_{qds} - \sigma\, L_s\, i_{qds})$$

$$V_{qdr} = 0 = i_{qdr}\, r_r + P\lambda_{qdr} - j\omega_r\, \lambda_{qdr} \qquad (4.13)$$

$$(r_r/L_m - j\omega_r\,(L_r/L_m))\,\lambda_{qds} + (L_r/L_m)\,P\lambda_{qds}$$

$$= (r_r L_s/L_m - j\omega_r\, L_r\, L_s\sigma/L_m)\, i_{qds} + L_r L_s\sigma/\, L_m\, P i_{qds} \qquad (4.14)$$

From equation (4.14), it is clear that the stator current can be expressed as a function of the stator flux linkage i.e.,

$$i_{qds} = F(\lambda_{qds}) \qquad (4.15)$$

Substituting in equation (4.12), we can conclude that

$$T_e = g(\lambda_{qds}) \qquad (4.16)$$

This together with the fact that the stator flux magnitude is constant makes the resulting equation of electromagnetic torque as a function of motor parameters, the constant stator flux modulus $|?_{qds}|$ and the stator flux position (s_s).

The proposed torque equation is

$$T_e = g\,(c.\exp\,(js_s)) \tag{4.17}$$

Equation (4.17) shows that the rate of change of the electromagnetic torque (dTe/dt) is proportional to the rate of change of the stator flux position (dσs/dt). Thus, a fast torque response can be obtained by controlling the stator flux position which in turn can be adjusted by selecting the appropriate stator flux vector. This can be achieved considering the stator voltage

$$V_{qds} = r_s\, i_{qds} + P?_{qds} \tag{4.18}$$

If we assume that the stator resistance voltage drop can be neglected then Equation 3.12 becomes

$$V_{qds} = P?_{qds} \tag{4.19}$$

4.5.1 Switching Strategy

From Equation 4.19 it can be seen that the inverter voltage directly force the stator flux, the required stator flux locus will be obtained by choosing the appropriate inverter switching state. Thus the stator flux linkage move in space in the direction of the stator voltage space vector at a speed that is proportional to the magnitude of the stator voltage space vector. The appropriate stator voltage vector is selected by means of step by step procedure, so as to change the stator flux as required.

If an increase of the torque is required then the torque is controlled by applying voltage vectors that advance the flux linkage space vector in the direction of rotation. If a decrease in torque is required then zero switching vector is applied, the zero vector that minimize inverter switching is selected.

For example if the stator flux vector lies in the k-th sector and the motor is running anticlockwise torque can be increased by applying stator voltage vectors V_{k+1} or V_{k+2}, and decreased by applying a zero voltage vector V_0 or V_7. Decoupled control of the torque and stator flux is achieved by acting on the radial and tangential components of the stator voltage space vector in the same directions, and thus can be controlled by the appropriate inverter switching.

In general, if the stator flux linkage vector lies in the k-th sector its magnitude can be increased by using switching vectors V_{k-1} (for clockwise rotation) or V_{k+1} (for anticlockwise rotation) and can be decreased by applying voltage vectors V_{k-2} (for clockwise rotation) or V_{k+2} (for anticlockwise rotation)

4.5.2 Lookup Table

The above can be tabulated in the look-up Table 3.1. The inputs to the look-up table are the torque error and the flux error generated by a three level hysteresis comparator and a two level hysteresis comparator respectively. The torque error is 1 if an increase in torque is required, 0 if a decrease is required.

For anticlockwise rotation:

$$dT = 1 \quad \text{if } T \le T_{ref\text{-}}|\Delta T| \tag{4.20 a}$$

$$dT = 0 \quad \text{if } T \ge T_{ref} \tag{4.20 b}$$

And for clockwise rotation:

$$dT = 1 \quad \text{if } |T| \le |T_{ref}| - |\Delta T| \tag{4.21 a}$$

$$dT = 0 \quad \text{if } T \le T_{ref} \tag{4.21 b}$$

Similarly, for an increase in flux, the flux error is 1 and for a decrease in flux, the flux error is 0.

That is:

$$d\lambda = 1 \quad \text{if } |\lambda| \le |\lambda_{ref}| - |\Delta\lambda| \tag{4.22 a}$$

$$d\lambda = 0 \quad \text{if } |\lambda| \ge |\lambda_{ref}| + |\Delta\lambda| \tag{4.22 b}$$

Table 2 Takahashi look-up table.

Flux Error $d\lambda$	Torque Error dT	S1	S2	S3	S4	S5	S6
1	1	V_2	V_3	V_4	V_5	V_6	V_1
	0	V_0	V_7	V_0	V_7	V_0	V_7
	-1	V_6	V_1	V_2	V_3	V_4	V_5
-1	1	V_3	V_4	V_5	V_6	V_1	V_2
	0	V_0	V_7	V_0	V_7	V_0	V_7
	-1	V_5	V_6	V_1	V_2	V_3	V_4

Table.3 Flux and Torque due to applied voltage
(Arrow indicate magnitude and direction)

Voltage vector	V_1	V_2	V_3	V_4	V_5	V_6	V_0 or V_7
ψ_s	↑	↑	↓	↓	↓	↑	0
Te	↓	↑	↑	↑	↓	↓	↓

4.6 ESTIMATION OF STATOR FLUX IN DTC

Accurate flux estimation in vector controlled induction motor drives is important to ensure proper drive operation and stability. The flux estimation techniques proposed is based on voltage model, current model, or the combination of both.

4.6.1 Current Model

The estimation based on current model described with the Equations (4.33-4.35) is normally applied at low frequency, and it requires the knowledge of the stator current and rotor mechanical speed or position.

$$\lambda_{qds} = ((r_r\,(L_m/L_r))/\,(P + \omega b_r))\; i_{qds} + \sigma L_s\, i_{qds} \qquad (4.33)$$

Where,

$$\omega b_r = r_r/L_r - j\omega_r \qquad (4.34)$$

$$\sigma = 1 - (L_m^{\,2}/\,L_sL_r) \qquad (4.35)$$

To get the accurate values of speed or position of the rotor, the encoders are used. The incremental use of encoder is undesirable because it reduces the robustness and reliability of the drive. The use of rotor parameters introduces error at high rotor speed due to the variation in rotor parameters. However it manages to eliminate the sensitivity due to variation in stator resistance.

4.6.2 Voltage Model

The voltage model, on the other hand, does not need a position sensor and the only motor parameter used is the stator resistance.

$$\lambda_{qs} = \int (V_{qs} - r_s\, i_{qs})\, dt \qquad (4.36\ a)$$

$$\lambda_{ds} = \int (V_{ds} - r_s\, i_{ds})\, dt \qquad (4.36\ b)$$

This gives accurate estimation at high speeds however, at low speed, some problems arise. In practical implementation, even a small DC off-set present in the back emf due to noise or measurement error inherently present in the current sensor, can cause the integrator to saturate. Hence it is not suitable for low speed.

4.6.3 Sliding Mode Observer

A sliding mode based observer can achieves robustness against stator resistance and speed variation without needing speed sensing or estimation. A state space model in the stationary reference frame is used in this observer; the stator currents and stator flux linkages are the state variables:

$$\begin{aligned} dx/dt &= Ax + Bu \\ y &= Cx \end{aligned} \tag{4.37}$$

Where,

$$x = [i_{qs} \;\; i_{ds} \;\; \lambda_{qs} \;\; \lambda_{ds}]^t$$

$$y = [i_{qs} \;\; i_{ds}]^t$$

$$u = [v_{qs} \;\; v_{ds}]^t$$

$$A = \begin{bmatrix} -((r_r/\sigma L_r) + (r_s / \sigma L_s)) & -\omega_r & r_r/\sigma L_s L_r & \omega_r / \sigma L_s \\ \omega_r & -((r_r/\sigma L_r) + (r_s / \sigma L_s)) & -\omega_r / \sigma L_s & r_r/\sigma L_s L_r \\ -r_s & 0 & 0 & 0 \\ 0 & -r_s & 0 & 0 \end{bmatrix}$$

$$B = \begin{bmatrix} 1/\sigma L_s & 0 \\ 0 & 1/\sigma L_s \\ 1 & 0 \\ 0 & 1 \end{bmatrix}$$

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

The matrices A and B contain the machine parameters and the rotor speed. Of all the parameters, the stator resistance and the rotor speed are the most uncertain and they are the main source of sluggish response of the DTC controller.

If the uncertainty in the values of other motor parameters is neglected, matrices A and B can be divided into two parts, one corresponding to nominal or constant parameters and the second to uncertain parameters, the system model can be rearranged in Equation 2.31 form.

$$\begin{aligned} \dot{x} &= \bar{A}x - \bar{B}u + \Delta Ax - \Delta Bu \\ y &= Cx \end{aligned} \tag{4.38}$$

Ā and B are the nominal state and input matrixes, which are supposed to be well known. ΔA and ΔB represent the uncertainty in A and B due to stator resistance and speed variations.

$$\Delta A = \begin{bmatrix} -\frac{\Delta r_s}{\sigma L_s} & -\Delta\omega_r & 0 & \frac{\Delta\omega_r}{\sigma L_s} \\ \Delta\omega_r & -\frac{\Delta r_s}{\sigma L_s} & -\frac{\Delta\omega_r}{\sigma L_s} & 0 \\ -\Delta r_s & 0 & 0 & 0 \\ 0 & -\Delta r_s & 0 & 0 \end{bmatrix}$$

$$\Delta B = 0$$

where Δr_s and Δw_r are bounded stator resistance and rotor speed uncertainties.

The observer tracks the stator currents components, by putting the corresponding currents errors into sliding mode, and thereby, imposes the asymptotic convergence of the flux observation errors despite parameter variations. Fig. 2.8 shows the general layout of the sliding mode observer.

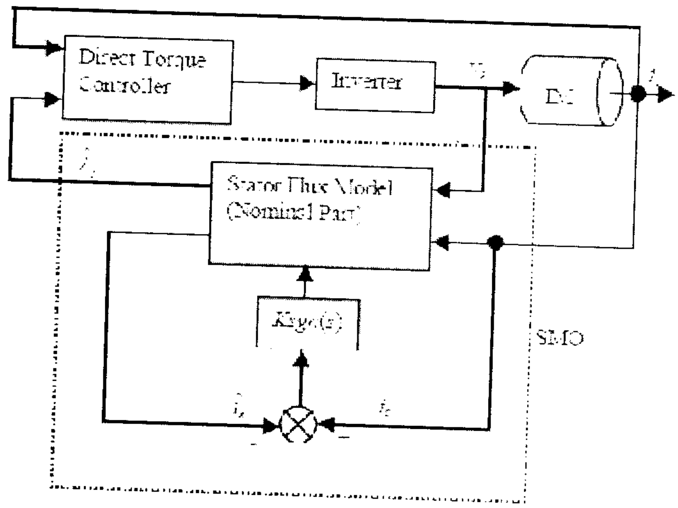


Fig 4.10 Sliding mode observer structure

The observer is based on the nominal part of the rearranged stator flux model with stator current and voltage measurements as inputs. The system is then split into two coupled subsystems: the first one corresponds to the measurable stator currents and the second to the stator flux components to be reconstructed.

$$X_1 = A_{11}y + A_{12}\,x_2 + B_1u + k_1\mathrm{sgn}(s)$$

$$X_2 = A_{21}y + u + k_2\mathrm{sgn}(s)$$

Where, $X_1 = [i_{qs}\ \ i_{ds}]^t$ and $X2 = [\lambda_{qs}\ \ \lambda_{ds}]^t$

$$\bar{A} = \begin{bmatrix} \bar{A}_{11} & \bar{A}_{12} \\ \bar{A}_{21} & 0 \end{bmatrix} \text{ and } \bar{B} = \begin{bmatrix} \bar{B}_1 \\ 0 \end{bmatrix}$$

The vector s is the current subsystem switching function; it is directly related to the current observation error. k_1 and k_2 are gain matrixes to be designed. The switching term sgn(s) acts as a control input to the observer structure. It is based on the sign of the current observation error instead of its actual value. This is what gives the sliding mode observer its well-known properties of robustness, stability, and good dynamic response.

4.7 OVERALL DIRECT TORQUE CONTROL MODEL

The DTC model consists of a three phase VSI fed inverter that performs the conversion process. The 3 phase AC supply is given to the induction motor for achieving the control. From the induction motor the stator current and voltage values are measured using the measurement block. From the measured values the torque and flux are calculated directly and then the inverter switching is done based on the voltage sector selection values and the respective voltage vectors as per the requirement from the preset switching table. The fig.4.11 shows the overall direct torque control block.

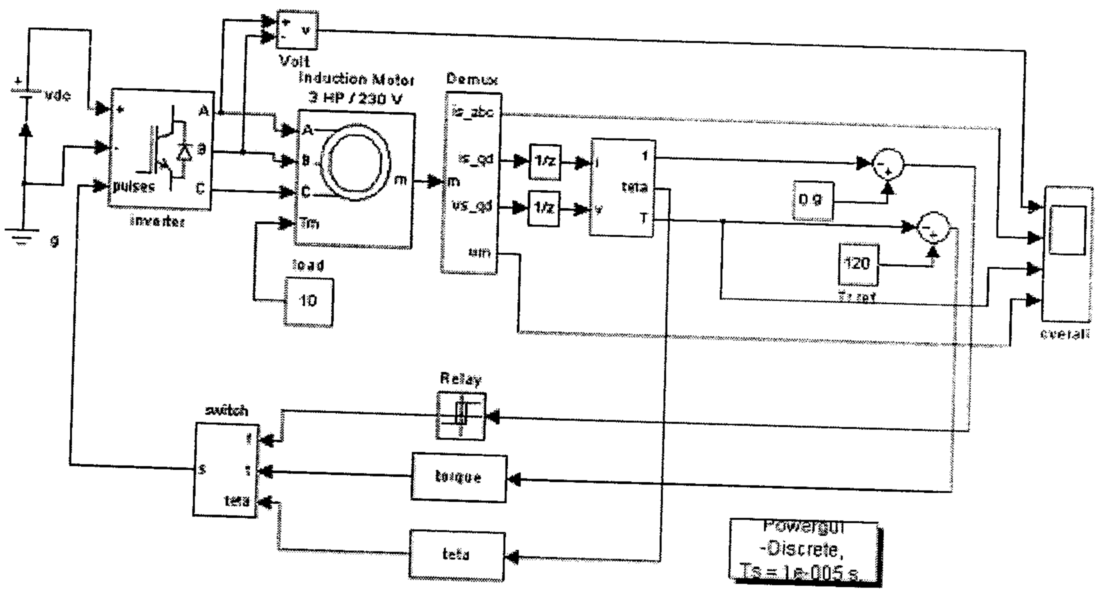


Fig.4.11 Shows the overall direct torque control block

4.7.1 Torque and Flux Calculator

Fig.4.6 shows the block diagram of the torque and flux calculator. The voltage and current values obtained from the DTC measure block acts as the input to the torque and flux calculator block. The voltage and current values are transformed in to their respective dq reference frames. Then, the torque, flux and angle values are obtained by means of transfer function method from the dq reference signals.

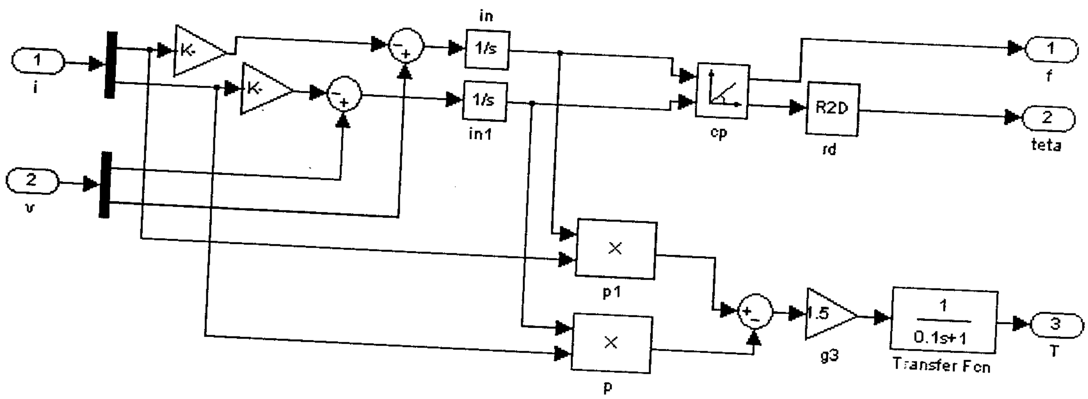


Fig. 4.12 Block diagram of Torque and Flux Calculator

The torque and the flux hysteresis controller block takes the reference torque and flux values along with the calculated torque and flux values from the torque and flux calculator. The information to control power switches is produced in the torque and flux controller block. Both the actual torque and actual flux are fed to the comparators where they are compared, every 25 microseconds, to the torque and flux reference values. Torque and flux status signals are then calculated using a two level hysteresis control methods. These signals are then fed to the optimum pulse selector or switching table.

It takes various input angles and checks for the condition through the logical and relational operators. Then it converts those outputs into the radians which are then summed and checked for saturation to produce the output sector values, with the help of these sector values the switching values is selected. The outputs from the torque and flux hysteresis controller and the flux sector selection acts as the input to the input to the switching table. The switching table is the block where the optimal vector for the inverter is decided and given as the gate to the inverter unit. The vectors are selected based on whether an increase or decrease in torque or flux is required.

CHAPTER V
EXPERIMENTAL SETUP

5.1 INTRODUCTION

The Direct Torque Control was implemented using a setup consisting of a six step voltage inverter and TMS320F240 digital signal processor (DSP) that controlled and generated the switching commands for the inverter's gate drive circuit.

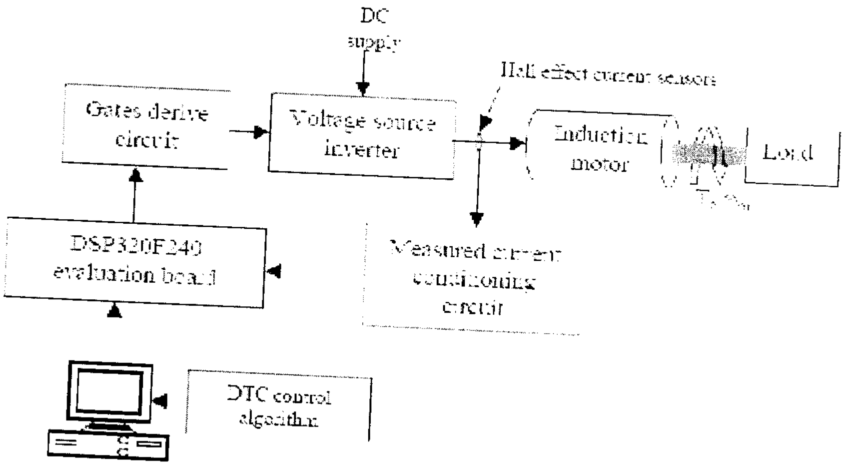


Fig. 5.1 Block diagram of the experimental setup

The induction motor current was measured using two Hall Effect current sensors and fed back to the processors on board analog-to-digital converter through a conditioning circuit. The block diagram of the experimental setup is shown in Fig. 5.1. The details of the setup components are discussed next.

5.2 VOLTAGE SOURCE INVERTER AND THE GATE DRIVE CIRCUIT

The details of the six step voltage inverter were discussed in Section 4.3. The topology of the inverter was shown in Fig.4.2. The co packs IGBT and diode "IRG4BC30FD" was used to build the inverter. For the IGBT to turn on, the gate voltage should be 10 to 15 V higher than the emitter voltage. For the high side switches, such gate voltage has to be higher than the DC bus high side rail, which is the highest voltage available in the drive circuit. The gate voltage must be controllable from the processor logic, which is normally referenced to ground; thus, the control signals have to be level-shifted to the emitter of the high side switch, which swings between the two rails of the DC bus. Fig. 5.2 shows the gate drive circuit used for the voltage source inverter. The IR2130, which has six-gate drivers, performs the required interfacing between the logic level control commands, and the high power IGBT switches, and it level-shifts the switching commands for the high side switches. A floating DC power supply (or bootstrap capacitor) was used for each of the three high side gate drivers where the emitter voltage swings between the two rails of the DC bus. The low side gate drivers were referenced to the DC bus low rail so they were directly fed with the control signals. The inputs of the IR2130 have inverted logic; the output goes high when the corresponding input command goes low. For isolation between the DSP and the power circuit, HCPL2531 high-speed transistor opto-couplers were used.

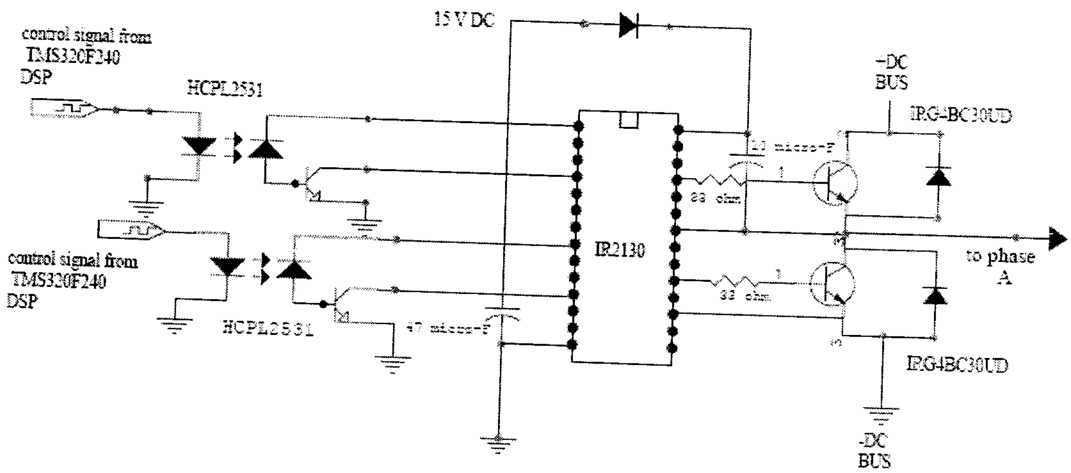


Fig. 5.2 The gate drive circuit

5.3 THE TMS320F240 EVALUATION BOARD

The TMS320F240 board is a 16-bit fixed point DSP that can execute 20 million instructions per second (MIPS). This DSP is specially designed for motor control applications. Included in the DSP event manager are special pulse-width modulation (PWM) generation functions that have 12 outputs with 0 - 5 V logic levels, a programmable dead-band function and a space vector PWM state machine for 3-phase motors that provides maximum efficiency in the switching of power transistors. Three independent up/down timers and three compare registers support the generation of asymmetric (non-centered) as well as symmetric (centered) PWM waveforms. Symmetrical PWM was used in the DTC control implementation. In this method the general purpose timer is set in the up-down continuous mode where its count register starts counting from zero up to the value stored in the timer period register, then it counts down to zero. The value in the timer period register is set to make the period of the up-down counting equal to the switching period of the drive. The value stored in the compare register is a function of the duty ratio and it decides how long the PWM output will be active. When the value of the timer count registers during up counting matches the value stored in the compare register, the PWM output is set active and it stays active until the value of the count register during down counting matches the compare register value. The PWM output is not active before the first match (during up counting) and after the second match (during down counting). Fig. 6.3 illustrates the PWM generation; the polarity of the PWM outputs is set to match the gate drive requirement. When a nonzero voltage vector is selected, the compare register is set to match the duty ratio value.

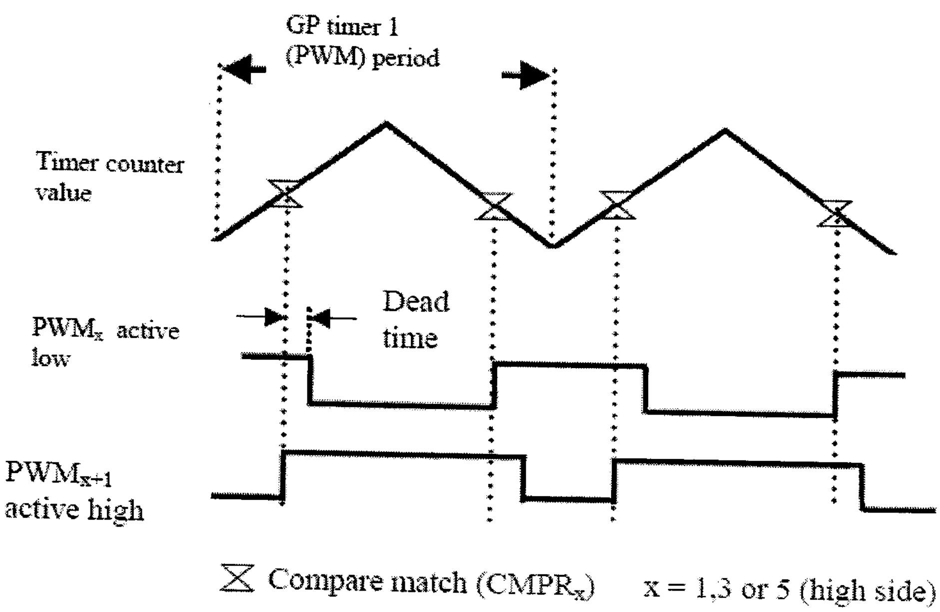


Fig. 5.3 Symmetrical PWM waveform generation

The programmable dead band feature of the PWM generator prevents shoot through by delaying the leading edge of the turn on pulse to the upper switches in order to allow sufficient time for a conducting switch to turn off before the other switch in the same phase turns on. The IRG4BC30FD IGBTs used in the inverter have a typical 91-ns turn off delay time. A dead band of 3 μs was programmed to ensure safe operation of the inverter. Also included on the DSP board is a four channel 12 bit digital to analog (D/A) converter that has an unsigned 0-5 V output range.

5.4 THE CURRENT SENSOR CIRCUIT

The DSP has a dual unsigned 10-bit analog-to-digital converter with internal sample-and-hold circuitry. The upper and lower reference voltages for the converter can be set to any value in the range 0-5 V. The unsigned digital result of the conversion process for the 10-bit A/D converter is presented by the following equation: Digital result = 1023 * (Input voltage/Reference voltage). Two of the induction motor phase currents (i_a, i_b) were measured using Hall effect sensors that have a turn ratio of 1:1000. The sensor's output current was converted to a voltage by placing a measuring resistance between the output of the secondary coil and the ground. A capacitor was added to attenuate high frequencies. Since the motor currents are bipolar, an offset of 2.5 V was added to the voltage across the measuring resistance to keep the input to the A/D converter within the 0-5 V A/D converter range. The current sensor circuit is shown in Fig. 5.4

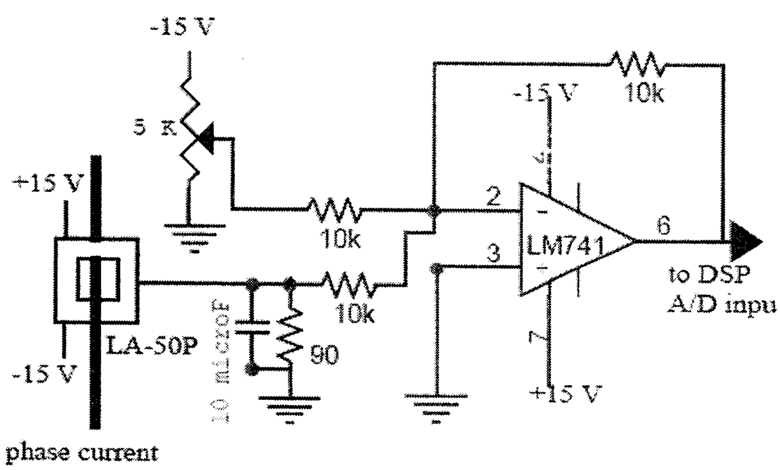


Fig. 5.4 The current sensor circuit

CHAPTER VI

SIMULATION RESULTS

6.1 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.

The vector control of the induction motor is developed using Simulink. The Induction Motor parameters are as follows

Table 4. Machine Parameters

Pn	Nominal power	3HP
Rs	Stator resistance	0.087Ω
Rr	Rotor resistance	0.228Ω
Ls	Stator inductance	0.008H
Lr	Rotor inductance	0.008H
Lm	Mutual inductance	34.7mH
J	Rotor inertia	1.662 kg.m2
2p	Number pole pairs	2

6.2 VECTOR CONTROLLER MODEL

The induction motor is fed by a current-controlled PWM inverter, which operates as a three-phase sinusoidal current source. The motor speed ω is compared to the reference ω* and the error is processed by the speed controller to produce a torque command Te*. The iqs* and ids* current references are converted into phase current references ia*, ib*, ic* for the current regulators. The simulation block diagram of Vector control method is shown in fig. 6.1.

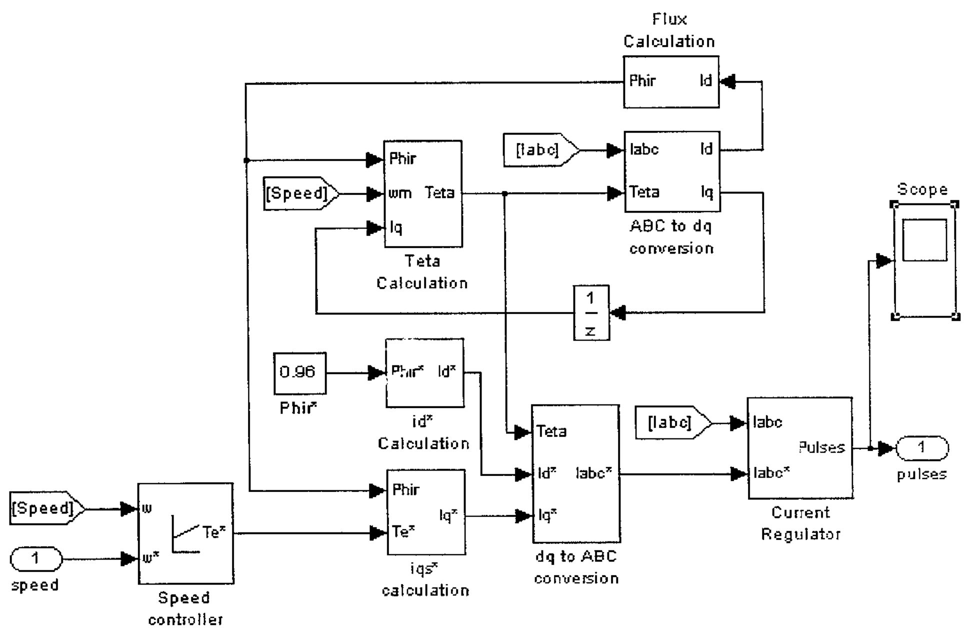


Fig. 6.1 Vector Control of induction motor

The regulators process the measured and reference currents to produce the inverter gating signals. The output wave forms of stator current, voltage, electromagnetic torque, speed are shown in fig. 6.2. The role of the speed controller is to keep the motor speed equal to the speed reference input in steady state and to provide a good dynamic during transients. It can be of proportional-integral type.

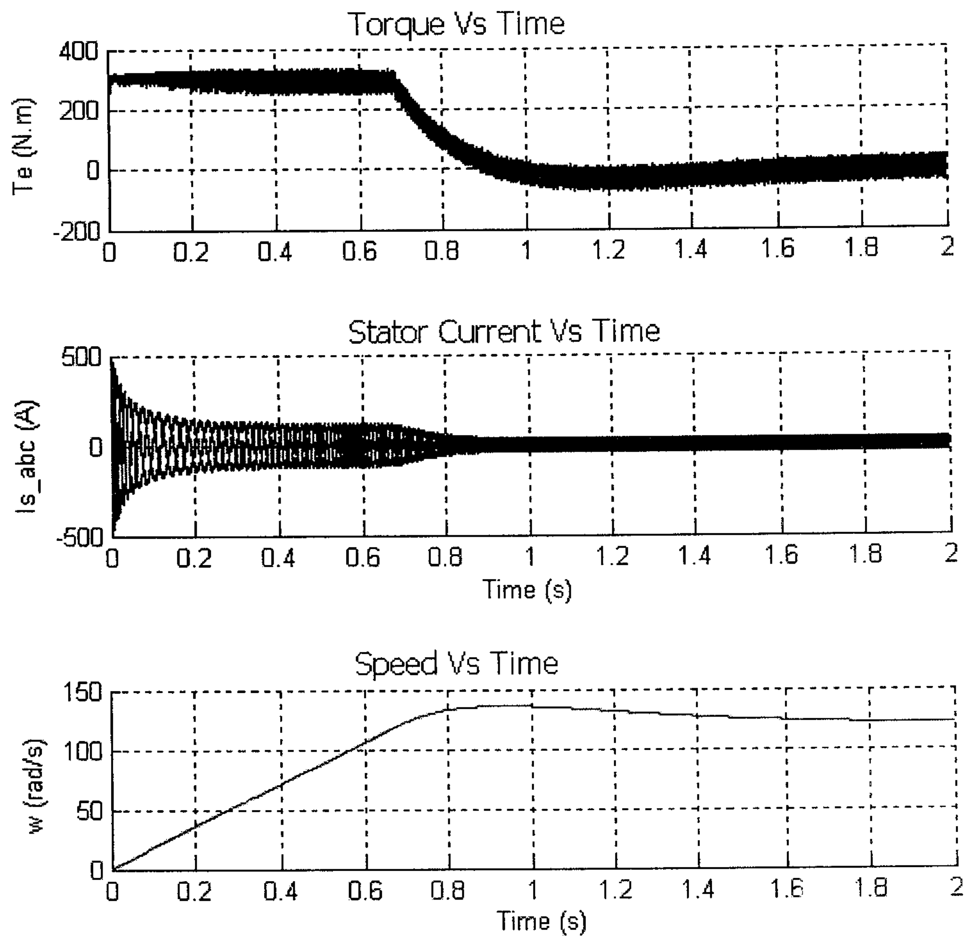


Fig. 6.2 (a) Stator Current in Amp (b) Torque in Nm (c) Speed in rad/s of the Vector control of Induction motor

6.3 TORQUE AND FLUX CONTROLLER

Fig.6.3 shows the block diagram of the torque and flux calculator. The voltage and current values obtained from the DTC measure block acts as the input to the torque and flux calculator block. The voltage and current values are transformed in to their respective dq reference frames. Then, the torque, flux and angle values are obtained by means of transfer function method from the dq reference signals. The controller takes the reference torque and flux values along with the calculated torque and flux values from the torque and flux calculator.

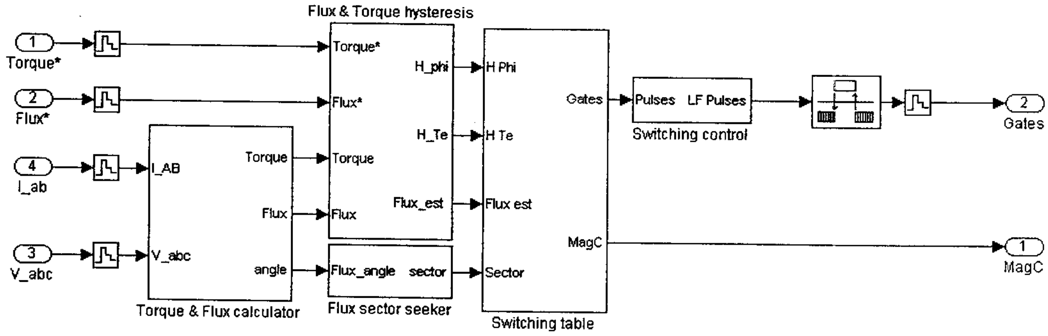
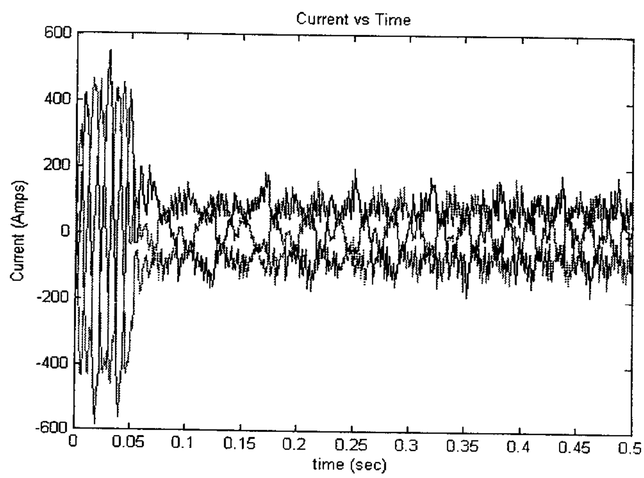
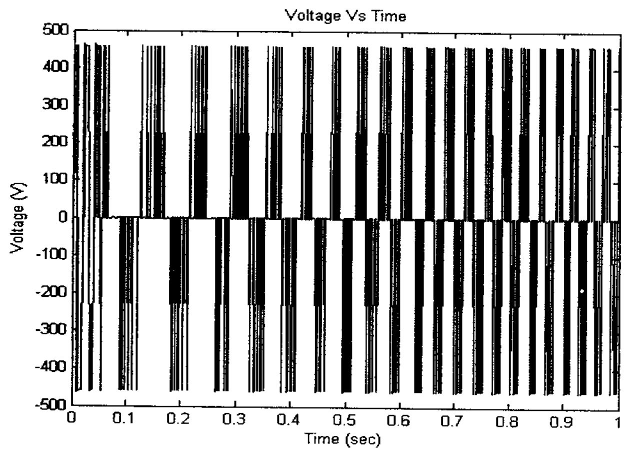


Fig. 6.3 Torque and Flux Controller

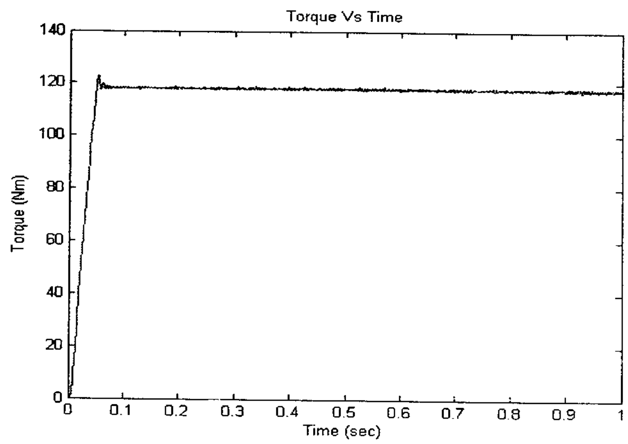
The information to control power switches is produced in the torque and flux controller block. Both the actual torque and actual flux are fed to the comparators where they are compared, every 25 microseconds, to the torque and flux reference values. Torque and flux status signals are then calculated using a two level hysteresis control methods. These signals are then fed to the optimum pulse selector or switching table.



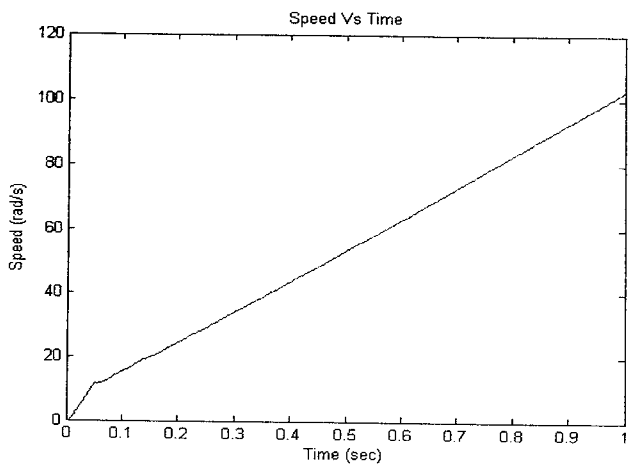
(A). Stator Current Vs Time



(B). Voltage Vs Time



(C). Electromagnetic Torque Vs Time



(D). Speed Vs Time

Fig. 6.4 Output Waveform of Direct Torque Controller

CHAPTER VII

CONCLUSION AND FUTURE WORK

7.1. CONCLUSION

Direct torque control is supposed to one of the best controllers for driving any induction motor. Its main principle have been introduced and deeply explained. It is also demonstrated in this thesis that the method of DTC also allows the independent and decoupled control of motor torque and motor stator flux.

Two different estimators for the stator flux and torque have been fully developed.

It is also apparent from the investigation reported that DTC strategy is simpler to implement than the flux vector control method because voltage modulators and co-ordinate transformations are not required.

The theoretical claim that the DTC induction motor drive has better dynamic response was verified by simulation using MATLAB. The use MATLAB gave satisfactory results and reduces the computation burden by avoiding unnecessary complex mathematical modeling of the nonlinear systems. By using MATLAB the specific motor performance can be achieved at a lower switching frequency compared to the vector control method, which in turn increases the efficiency of the drive by reducing losses due to currents and flux harmonics.

7.2 FUTURE WORK

All future work is summarized schematically in the following ideas:

- ➢ Development of new Fuzzy Logic Controllers to achieve better performance. The 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 appears in any power drives.
- ➢ DTC with Duty Ratio Control.
- ➢ DTC with Multi Level Converter.

APPENDIX

CODING FOR MATLAB SIMULATION

S function for Torque calculation:

```
function [sys,x0,str,ts]=torque(t,x,u,flag)
switch flag
case 0
[sys,x0,str,ts] = mdlInitializeSizes;
case 2
sys = [];
case {1,9}
sys = [];
case 3
sys = mdlOutputs(t,x,u);
otherwise
error(['unhandled flag = ',num2str(flag)]);
end
function [sys,x0,str,ts] = mdlInitializeSizes
sizes= simsizes;
sizes.NumContStates  =0;
sizes.NumDiscStates  = 0;
sizes.NumOutputs     = 1;
sizes.NumInputs      = 1;
sizes.DirFeedthrough = 1;
sizes.NumSampleTimes = 1;
sys = simsizes(sizes);
str = [];
x0  = [];
ts  = [1e-5 0];
```

```
function sys = mdlOutputs(t,x,u)
if(u<-2)
sys=-1;
elseif((-2<u)&(u<2))
sys=0;
else
sys=1;
end
```

/* To get output [illegible]
/* using dT=1 [illegible]
dT=0 [illegible]

S function for Flux calculation:

```
function [sys,x0,str,ts]=teta(t,x,u,flag)
switch flag
case 0
[sys,x0,str,ts] = mdlInitializeSizes;
case 2
sys = [];
case {1,9}
sys = [];
case 3
sys = mdlOutputs(t,x,u);
otherwise
error(['unhandled flag = ',num2str(flag)]);
end
function [sys,x0,str,ts] = mdlInitializeSizes
sizes= simsizes;
sizes.NumContStates  =0;
sizes.NumDiscStates  = 0;
sizes.NumOutputs     = 1;
sizes.NumInputs      = 1;
sizes.DirFeedthrough = 1;
```

/* INPUT teta variable declaration *

```
sizes.NumSampleTimes = 1;
sys = simsizes(sizes);
str = [];
x0  = [];
ts  = [1e-5 0];                                   /* setting sampling time */
function sys = mdlOutputs(t,x,u)
if((0<u)&(u<=60))
sys=2;
elseif((60<u)&(u<=120))
sys=1;
elseif((120<u)&(u<=180))
sys=6;
elseif((-180<u )&(u<=-120))
sys=5;
elseif((-120<u)&(u<=-60))
sys=4;
else
sys=3;
end
```

S function for Switching Table:

```
function [sys,x0,str,ts] = switching(t,x,u,flag)  /* switching table function
switch flag,                                          declaration */
case 0,
[sys,x0,str,ts] = mdlInitializeSizes;
case 2,
sys = [];
case 3,
sys = mdlOutputs(t,x,u);
case 9,
```

```
sys = [];
otherwise
error(['unhandled flag = ',num2str(flag)]);
end
function [sys,x0,str,ts] = mdlInitializeSizes

sizes = simsizes;
sizes.NumContStates  = 0;
sizes.NumDiscStates  = 0;
sizes.NumOutputs     = 1;
sizes.NumInputs      = 3;        /* Torque error flux error angle as inputs */
sizes.DirFeedthrough = 1;
sizes.NumSampleTimes = 1;
sys = simsizes(sizes);
x0  = [];
str = [];
ts  = [1e-5 0];
function sys = mdlOutputs(t,x,u)      /* checking for the variable based on switching table */
if(u(1)== 1 & u(2)==1&u(3)==1)
sys=6;
elseif(u(1)== 1 & u(2)== 0&u(3)==1)
sys=8;
elseif(u(1)== 1 & u(2)==-1&u(3)==1)
sys=2;
elseif(u(1)== 0 & u(2)== 1&u(3)==1)
sys=5;
elseif(u(1)== 0 & u(2)== 0&u(3)==1)
sys=7;
elseif(u(1)== 0 & u(2)==-1&u(3)==1)
sys=3;
elseif(u(1)== 1 & u(2)==1&u(3)==2)
```

```
sys=1;
elseif(u(1)== 1 & u(2)== 0&u(3)==2)
sys=7;
elseif(u(1)== 1 & u(2)==-1&u(3)==2)
sys=3;
elseif(u(1)== 0 & u(2)== 1&u(3)==2)
sys=6;
elseif(u(1)== 0 & u(2)== 0&u(3)==2)
sys=8;
elseif(u(1)== 0 & u(2)==-1&u(3)==2)
sys=4;
elseif(u(1)== 1 & u(2)==1&u(3)==3)
sys=2;
elseif(u(1)== 1 & u(2)== 0&u(3)==3)
sys=8;
elseif(u(1)== 1 & u(2)==-1&u(3)==3)
sys=4;
elseif(u(1)== 0 & u(2)== 1&u(3)==3)
sys=1;
elseif(u(1)== 0 & u(2)== 0&u(3)==3)
sys=7;
elseif(u(1)== 0 & u(2)==-1&u(3)==3)
sys=5;
elseif(u(1)== 1 & u(2)==1&u(3)==4)
sys=3;
elseif(u(1)== 1 & u(2)== 0&u(3)==4)
sys=7;
elseif(u(1)== 1 & u(2)==-1&u(3)==4)
sys=5;
elseif(u(1)== 0 & u(2)== 1&u(3)==4)
sys=2;
```

```
elseif(u(1)== 0 & u(2)== 0&u(3)==4)
sys=8;
elseif(u(1)== 0 & u(2)==-1&u(3)==4)
sys=6;
elseif(u(1)== 1 & u(2)==1&u(3)==5)
sys=4;
elseif(u(1)== 1 & u(2)== 0&u(3)==5)
sys=7;
elseif(u(1)== 1 & u(2)==-1&u(3)==5)
sys=6;
elseif(u(1)== 0 & u(2)== 1&u(3)==5)
sys=3;
elseif(u(1)== 0 & u(2)== 0&u(3)==5)
sys=8;
elseif(u(1)== 0 & u(2)==-1&u(3)==5)
sys=1;
elseif(u(1)== 1 & u(2)==1&u(3)==6)
sys=5;
elseif(u(1)== 1 & u(2)== 0&u(3)==6)
sys=7;
elseif(u(1)== 1 & u(2)==-1&u(3)==6)
sys=1;
elseif(u(1)== 0 & u(2)== 1&u(3)==6)
sys=4;
elseif(u(1)== 0 & u(2)== 0&u(3)==6)
sys=8;
else
sys=2;
end
```

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