3.1 Introduction

In the past, DC motors were extensively used in applications where high performance variable speed operation and controlled torque were required. The flux and torque can be easily controlled by armature and field current in DC motors. This reduces the complexity of speed and torque control applications. The major drawback of DC motor is the use of commutator and brushes. Also DC motors require regular maintenance and limited use in explosive environments as well as in high speed and high voltage applications. AC Motors are now replacing DC motors in variable speed applications. Among the AC motors, IMs are cost effective, maintenance free and are suitable for variable speed applications in terms of size, weight, speed of rotation, efficiency, controllability and reliability. The variable speed drives are used in all industries to control the speed of IM driving loads ranging from pumps and fans to complex drives in paper machines, rolling machines, cranes and similar drives. Due to the high demand for flexibility in manufacturing and due to the urge for rational use of electrical energy, the requirement of adjustable speed drives is growing in industrial applications. The important factors to consider in the adjustable speed drives are

a. Constant Torque: Maintain torque constant irrespective of speed variation

b. Variable Torque: Torque should be variable such as low torque at low speeds and high torque at high speeds.
c. Constant Power: Maintain high torque at low speeds and low torque at high speeds

The other factors need to be considered while choosing adjustable speed drives for speed, torque and position control application are rating, cost, speed range, efficiency, speed regulation, braking requirements, reliability, power factor, power supply availability and environmental considerations [51]. AC motor mathematical models are more complex than those of DC motor. Hence more complex control schemes are required to achieve speed and torque control. In the last three decades, high performance control strategies for IMs are developed to meet the industry requirements. It is now recognized that the two high performance control strategies for IM drives are FOC [52 - 55] developed in 1970’s and DTC [56 - 60] developed in 1980’s. These control strategies control the torque and flux of the motor effectively and force the motor to track the reference commands effectively regardless of the parameter variations occurring in motor and/or load. Many researchers focussed FOC and DTC techniques over the last 30 years and found that FOC is dependent on knowledge of rotor parameters such as time constant. The traditional DTC technique is a non constant converter switching frequency which may result in high losses. Both control strategies have their own advantages and disadvantages and they are implemented successfully in industrial applications. In this chapter the principle and performance of FOC and DTC techniques for IM using MC drive are presented. The d-q model of IM used in FOC and DTC techniques is also discussed.
3.2 d-q model of IM

Fig. 3.1 shows the induction machine d-q equivalent circuit. From the equivalent circuit the modelling equations were obtained.

\[
\frac{df_{qs}}{dt} = \omega_b \left[V_{qs} - \frac{\omega_e}{\omega_b} F_{ds} + \frac{R_s}{x_{ls}} (F_{mq} + F_{ds})\right]
\]  
(3.1)

\[
\frac{df_{ds}}{dt} = \omega_b \left[V_{qs} + \frac{\omega_e}{\omega_b} F_{qs} + \frac{R_s}{x_{ls}} (F_{md} + F_{ds})\right]
\]  
(3.2)

\[
\frac{df_{qr}}{dt} = \omega_b \left[V_{qr} - \frac{(\omega_e - \omega_r)}{\omega_b} F_{dr} + \frac{R_r}{x_{lr}} (F_{mq} + F_{qr})\right]
\]  
(3.3)

\[
\frac{df_{dr}}{dt} = \omega_b \left[V_{dr} + \frac{(\omega_e - \omega_r)}{\omega_b} F_{qr} + \frac{R_r}{x_{lr}} (F_{mq} + F_{dr})\right]
\]  
(3.4)

\[
F_{mq} = x^*_m \left[\frac{f_{qs}}{x_{ls}} + \frac{f_{qr}}{x_{lr}}\right]
\]  
(3.5)

\[
F_{md} = x^*_m \left[\frac{f_{ds}}{x_{ls}} + \frac{f_{dr}}{x_{lr}}\right]
\]  
(3.6)

\[
i_{qs} = \frac{1}{x_{ls}} (F_{qs} - F_{mq})
\]  
(3.7)

\[
i_{ds} = \frac{1}{x_{ls}} (F_{ds} - F_{md})
\]  
(3.8)
An IM can be represented with five equations. To solve these equations, they need to be in the form of state space. This was achieved by substituting equations \( e \) and \( f \) into \( a-d \) and grouping the similar terms so that each state derivative is a function of state variables and inputs (the details of the work are presented in [95]). The resulting equations are given below [51]. The d-q model of IM is shown in Fig.3.2.

\[
id_q = \frac{1}{x_{lr}} (F_{q} - F_{m_q}) \quad (3.9)
\]

\[
id_d = \frac{1}{x_{lr}} (F_{d} - F_{m_d}) \quad (3.10)
\]

\[
T_e = \frac{3}{2} \left( \frac{p}{2} \right) \frac{1}{\omega_b} (F_{ds} i_{qs} - F_{qs} i_{ds}) \quad (3.11)
\]

\[
T_e - T_L = f \left( \frac{2}{p} \right) \frac{d\omega_r}{dt} \quad (3.12)
\]

where \( d \): direct axis,

\( q \): quadrature axis,

\( s \): stator variable,

\( r \): rotor variable,

\( F_{ij} \): flux linkage \((i=q \text{ or } d \text{ and } j=s \text{ or } r)\),

\( v_{qs}, v_{ds} \): \( q \) and \( d \)-axis stator voltages,

\( v_{qr}, v_{dr} \): \( q \) and \( d \)-axis rotor voltages,

\( F_{mqs}, F_{md} \): \( q \) and \( d \)-axis magnetising flux linkages,

\( R_r \): rotor resistance,
$R_s$: stator resistance,

$X_{ls}$: stator leakage reactance ($\omega X_{Ls}$),

$X_r$: rotor leakage reactance ($\omega X_{Lr}$),

$X_{lm}^*$: $1/(\frac{1}{x_m} + \frac{1}{x_{ls}} + \frac{1}{x_{lr}})$,

$i_{qs}, i_{ds}$: $q$ and $d$-axis stator currents,

$i_{qr}, i_{dr}$: $q$ and $d$-axis rotor currents,

$p$: number of poles,

$J$: moment of inertia,

$T_e$: electrical output torque,

$\omega_e$: stator angular electrical frequency,

$\omega_m$: motor angular electrical base frequency,

$\omega_r$: rotor angular electrical speed,

### 3.3 Field Oriented Control System
#### 3.3.1 Clarke Transformation ($a, b, c$ to $\alpha, \beta$) Model

To achieve closed loop control, three-phase stator currents need to be measured and be expressed in two-phase system. Forward Clarke transformation converts system from abc (3-phase) to $\alpha\beta$ (2-phase). The vectorial representation of the space vectors and their projections are shown in Fig.3.3.

If we assume that $a$- and $\alpha$-axis are in the same direction, the 2-phase currents can be represented in terms of 3-phase currents as follows:

$$i_{sa} = k \left[ i_{sa} - \frac{1}{2} i_{sb} - \frac{1}{2} i_{sc} \right] \quad (3.18)$$

$$i_{sb} = k \frac{\sqrt{3}}{2} (i_{sb} - i_{sc}) \quad (3.19)$$

where:

$i_{sa}, i_{sb}, i_{sc}$: phase $a$, $b$, $c$ motor currents.
Fig. 3.2: d-q model of IM
For non-power-invariant transformation the constant $k$ equals $2/3$. This means $i_{sa}$ and $i_a$ are equal. Then the two-phase currents can be written as:

$$i_{sa} = i_{sa}$$

(3.20)

$$i_{s\beta} = \frac{1}{\sqrt{3}}i_{sa} + \frac{2}{\sqrt{3}}i_{sb}$$

(3.21)

### 3.3.2 Rotor Flux Calculator Model

After obtaining 2-phase representation of stator current, rotor flux and angle must be estimated. The rotor flux space vector magnitude and position is the most important information in FOC vector control of IM. The rotor magnetic flux space vector is used to calculate the rotating coordinates system (d-q). The flux model implemented here uses monitored rotor speed and stator currents. The error in the calculated value of the rotor flux, influenced by changes in temperature, is negligible for this rotor model [62] [63]. The flux model equations are derived from IM model and given as follows [64]:

$$[(1 - \sigma)Ts + Tr] \frac{d\psi_{\alpha}}{dt} = \frac{L_m}{R_s} u_{sa} - \psi_{\alpha} - \omega_r T_r \psi_{r\beta} - \sigma L_m T_s \frac{di_{sa}}{dt}$$

(3.22)

$$[(1 - \sigma)Ts + Tr] \frac{d\psi_{\beta}}{dt} = \frac{L_m}{R_s} u_{sb} - \psi_{\beta} - \omega_r T_r \psi_{r\beta} - \sigma L_m T_s \frac{di_{sb}}{dt}$$

(3.23)
\[ \psi_{rd} = \sqrt{\psi_{r\alpha}^2 + \psi_{r\beta}^2} \] 

(3.24)

\[ \sin \theta_{field} = \frac{\psi_{r\beta}}{\psi_{rd}} \] 

(3.25)

\[ \cos \theta_{field} = \frac{\psi_{r\alpha}}{\psi_{rd}} \] 

(3.26)

The transformation from the d-q to the αβ coordinate system is done using \(\sin \theta_{field}\) and \(\cos \theta_{field}\) values obtained here. The flux value estimated here is used as feedback for flux PI controller.

where:

\[ T_r = \frac{L_r}{R_r}\] is rotor time constant [s]

\[ T_s = \frac{L_s}{R_s}\] is stator time constant [s]

\[ \sigma = 1 - \frac{L_m}{L_d} \] is resultant leakage constant

\(u_{s\alpha}, u_{s\beta}, i_{s\alpha}, i_{s\beta}, \psi_{r\alpha}, \psi_{r\beta}\) are \(\alpha, \beta\) components of stator voltage, current and rotor flux space vectors respectively.

### 3.3.3 Park Transformation (α, β to d-q and backwards) Model

After performing forward Clarke transformation the outputs are used to get dq representation of stator currents. This is done using forward Park Transformation. For the d-axis aligned with the rotor flux, the Park Transformation can be presented vectorially as shown in Fig.3.4.

The d-q components are calculated using two-phase current from forward Clarke transformation block and angle from rotor flux calculator block. The expressions are as follows:

\[ i_{sd} = i_{sa} \cos \theta_{field} + i_{sb} \sin \theta_{field} \] 

(3.27)

\[ i_{sq} = -i_{sa} \sin \theta_{field} + i_{sb} \cos \theta_{field} \] 

(3.28)
The transformation from the d-q to the \( \alpha, \beta \) coordinate system is found by the following equations:

\[
\begin{align*}
    i_{sd} &= i_{sd} \cos \theta_{field} - i_{sq} \cos \theta_{field} \quad (3.29) \\
    i_{sb} &= i_{sd} \sin \theta_{field} + i_{sq} \cos \theta_{field} \quad (3.30)
\end{align*}
\]

This stage is needed after decoupling block. It uses the angle from the rotor flux calculator. The same rotor calculator uses the inverse Park’s transformation output voltages to perform its task.

### 3.3.4 Speed Controller Design for IM

The FOC control strategy in accordance with the rotor flux vector leads to simpler equations than those of d-axis aligned on the stator flux vector. The development of the FOC equation in accordance with the rotor flux vector results in expression of the motor torque being reduced to:

\[
T_{em} = p \frac{l_m}{l_r} \psi_r i_{sd} \quad (3.31)
\]

where \( p \) is the number of pole pairs and \( \psi_r \) is the rotor flux that can be calculated by:

\[
\frac{d\psi_r}{dt} = -\frac{R_r}{l_r} \psi_r + \frac{R_r M}{l_r} i_{sd} \quad (3.32)
\]

The motor mechanical equation that relates torque to speed is given by:
\[ T_m - T_L = J \frac{d\omega_r}{dt} \]  \hspace{1cm} (3.33)

where, \( J \) is the inertia constant.

In order to control the speed of the motor the following block diagram can be used:

![Speed Control Loop Diagram](image)

**Fig.3.5 Speed Control Loop [62]**

where

\[ \sigma = 1 - \frac{L_m^2}{L_s L_r} \]

\[ k = \frac{L_m}{L_r} \psi_r \]

The plant equation \( P(s) \) is the open loop transfer function of the system without the speed controller block shown in Fig.3.5. The \( P(s) \) can be derived under two conditions.

1. When load torque \( T_L = 0 \)

\[
P(s) = \frac{\omega(s)}{\omega_{ref}(s)} = \left( \frac{k}{(\sigma L_s/R_s)s + 1} \right) \left( \frac{1}{Js} \right) \]  \hspace{1cm} (3.34a)

2. When reference speed \( \omega_{ref} = 0 \)

\[
P(s) = \frac{T_e(s)}{T_L(s)} = \left( \frac{k}{(\sigma L_s/R_s)s + 1} \right) \left( \frac{1}{Js} \right) \]  \hspace{1cm} (3.34b)

The open-loop transfer function \( L(s) \) of the system with speed controller is given by

\[
L(s) = \left( \frac{k_p s + k_i}{s} \right) \left( \frac{k}{(\sigma L_s/R_s)s + 1} \right) \left( \frac{1}{Js} \right) \]  \hspace{1cm} (3.35)
3.3.5 Current Controller Design for Decoupled Control of IM

In order to perform FOC control, the plant equations or the transfer function in d- and q-axis are developed. The inputs to these functions are d-axis stator voltage \(V_{sd}\) and q-axis stator voltage \(V_{sq}\), and the outputs are d-axis stator current \(i_{sd}\) and q-axis stator current \(i_{sq}\).

(A). Transfer function for d-axis control

It is possible to control rotor flux by controlling stator current that is on the d-axis. This can be shown from the following equation.

\[
\frac{d\psi_r}{dt} = -\frac{R_r}{L_r} \psi_r + \frac{R_r M}{L_r} i_{sd} \tag{3.36}
\]

The stator voltage in the d-axis is given by

\[
V_{sd} = R_s i_{sd} + \sigma L_s \frac{di_{sd}}{dt} + \frac{M}{L_r} \frac{d\psi_r}{dt} - \omega_s \psi_{sq} \tag{3.37}
\]

By substituting

\[
i_{rq} = -\frac{M}{L_r} i_{sq} \tag{3.38}
\]

into the equation of stator flux

\[
\psi_{sq} = L_s i_{sq} + M i_{rq} \tag{3.39}
\]

we get stator flux expression to be

\[
\psi_{sq} = L_s i_{sq} - \frac{M^2}{L_r} i_{sd} = \left( L_s - \frac{M^2}{L_r} \right) i_{sq} = L_s \left( 1 - \frac{M^2}{L_r L_s} \right) i_{sq} = L_s \sigma i_{sq} \tag{3.40}
\]

By substituting it to the d-axis stator voltage, we get

\[
V_{sd} = R_s i_{sd} + \sigma L_s \frac{di_{sd}}{dt} + \frac{M}{L_r} \frac{d\psi_r}{dt} - \omega_s L_s \sigma i_{sq} \tag{3.41}
\]

Substituting the expression for \(\frac{d\psi_r}{dt}\) into the equation we get

\[
V_{sd} = R_s i_{sd} + \sigma L_s \frac{di_{sd}}{dt} + E_d \tag{3.42}
\]
where \( R_{sr} = R_s + R_r \left( \frac{M}{L_r} \right)^2 \), and the electrical force \( E_d = -R_r \frac{M}{L_r} \psi_r - \omega_s L_s \sigma i_{sq} \) represents the coupling between the d and q axis because of the presence of the term \( i_{sq} \).

(B). Transfer Function for q-axis control

The equation for q-axis stator voltage is given by:

\[
V_{sq} = R_s i_{sq} + \sigma L_s \frac{d i_{sq}}{dt} + \omega_s \psi_{sd} \tag{3.43}
\]

The d-axis stator flux is given by the following equation:

\[
\psi_{sd} = \sigma L_s i_{sd} + \frac{M}{L_r} \psi_r \tag{3.44}
\]

Substituting it into q-axis stator voltage we get:

\[
V_{sq} = R_s i_{sq} + \sigma L_s \frac{d i_{sq}}{dt} + \omega_s \frac{M}{L_r} \psi_r + \omega_s L_s \sigma i_{sd} \tag{3.45}
\]

Since one of rotor electrical equations is given by

\[
R_r i_{rq} + \omega_r \psi_r = 0, \tag{3.46}
\]

the rotor flux can be written in terms of q-axis stator current by substituting the expression for \( i_{rq} \) into the rotor electrical equation. Hence,

\[
\psi_r = R_r \frac{M}{\omega_r L_r} i_{sq} \tag{3.47}
\]

Substituting this equation to q-axis stator voltage we get:

\[
V_{sq} = R_s i_{sq} + \sigma L_s \frac{d i_{sq}}{dt} + \omega_s \left( \frac{M}{L_r} \right)^2 R_r i_{sq} + \omega_s L_s \sigma i_{sd} \tag{3.48}
\]

Therefore, \( V_{sq} \) can also be written as:

\[
V_{sq} = R_s i_{sq} + \sigma L_s \frac{d i_{sq}}{dt} + E_q \tag{3.49}
\]

where the electrical force \( E_q = \omega_s \sigma L_s i_{sd} + \omega_s \left( \frac{M}{L_r} \right)^2 R_r i_{sq} \), which represents the coupling between d-axis and q-axis because of the presence of the term \( i_{sd} \). The two axes d and q are decoupled by estimating the electric forces \( E_d \) and \( E_q \) as follows:


\[ E_d^e = -R_r \frac{M}{L_{r}} \psi_r^e - \omega_{s}^e L_{s} \sigma_{sq}^m \quad \text{and} \quad E_q^e = \omega_{s}^e \sigma L_{s} i_{sd}^m + \omega^m \left( \frac{M}{L_{r}} \right)^2 R_r i_{sq}^m \]. The term \( e \) represents the estimated variables, and the term \( m \) represents the measured variables.

These transfer functions can be represented as shown in Fig.3.6a and Fig. 3.6b. They describe the transfer functions of the IM if the d-axis is aligned on the rotor flux vector.

![Fig.3.6a: Transfer Function of IM in d-axis [62]](image1)

![Fig.3.6b Transfer Function of IM in q-axis [62]](image2)

The voltage to current IM plant equation is given by

\[ P(s) = \left( \frac{1/\sigma L_s}{s + R_{sr}/\sigma L_s} \right) \] \hspace{1cm} (3.50)

To test for step response, the plant equation must be written as

\[ P(s) = \left( \frac{1/\sigma L_s}{s + R_{sr}/\sigma L_s} \right) \left( \frac{1}{s} \right) \] \hspace{1cm} (3.51)

The open-loop transfer function is then given by

\[ L(s) = \left( \frac{k_p s + k_i}{s} \right) \left( \frac{1/\sigma L_s}{s + R_{sr}/\sigma L_s} \right) \] \hspace{1cm} (3.52)
When the q-axis $R_{sr}=R_s$, the plant equation is given by

$$P(s) = \left(\frac{1/\sigma L_s}{s + R_s/\sigma L_s}\right)$$  \hspace{1cm} (3.53)

and then the open-loop transfer function is given by

$$L(s) = \left(\frac{k_p s + k_i}{s}\right) \left(\frac{1/\sigma L_s}{s + R_s/\sigma L_s}\right)$$ \hspace{1cm} (3.54)

(C). Flux Controller

Flux control plant equation can also be obtained from the block diagram in Fig.3.9.

Taking into consideration the inner current control loop, it is found to be

$$P(s) = \frac{L_m R_f}{L_r} \frac{1}{s + \frac{R_f}{L_r}}$$ \hspace{1cm} (3.55)

3.3.6 Decoupled Control Model

To achieve rotor flux-orientated vector control, the direct axis stator current and the quadrature axis stator current responsible for producing rotor flux and torque respectively must be controlled independently. However, the stator voltage components equations are coupled. Hence, the direct axis component $u_{ds}$ and the quadrature component $u_{sq}$ both depend on both $i_{sq}$ and $i_{sd}$. These stator voltage components therefore cannot be considered as coupled control variables for rotor flux and electromagnetic torque [65 - 66]. However, the stator current components $i_{sd}$ and $i_{sq}$ can be controlled independently (decoupled control) provided the stator voltage are decoupled and the stator current components $i_{sd}$ and $i_{sq}$ are indirectly controlled by controlling stator terminal voltages of the IM.
The d-axis is rotating with the rotor flux vector, the d-axis and q-axis are decoupled by estimating the electric forces $E_d$ and $E_q$ as follows [68]:

$$ E_d^e = -R_r \frac{M}{L_r} \Psi_r^e - \omega_s^e L_s \sigma i_{sq}^m $$ (3.56)

and

$$ E_q^e = \omega_s^e \sigma L_s i_{sd}^m + \omega^m \left( \frac{M}{L_r} \right) R_r i_{sq}^m $$ (3.57)

The superscripts e and m denote the estimated and measured variables respectively. $\Psi_r^e$ is calculated by solving numerically the rotor flux equation. Its value is also used as a feedback for the rotor flux control closed loop.

$$ \frac{d\Psi_r}{dt} = - \frac{R_r}{L_r} \Psi_r + R_r \frac{M}{L_r} i_{sd} $$ (3.58)

$\omega_s^e$ is calculated from the following equation:

$$ \omega_s^e = \omega^m + R_r \frac{M}{\Psi_r} i_{sq}^m \omega^m = p \Omega^m = p \frac{d\theta}{td} $$ (3.59)

where, $\omega_s^e$, is the speed of the motor that can be measured using speed sensor, and $p$ is the number of pole pairs.

For an IM, $L_r/R_r$ is ten times bigger than $\sigma L_s/R_{sr}$, therefore it is possible to perform separation of poles by doing an inner closed loop for current and the outer closeloop for rotor flux [68].

Fig.3.7 shows that the d-axis closed loops are controlling the amplitude of the rotor flux, and Fig.3.8 shows that the closed loop of the q-axis is for controlling the stator current, which results in controlling the motor torque.
Thus the control algorithm works as follows: the three-phase currents are measured; they are then transformed by Park transformation to two-phase currents to calculate rotor angle. Then control variables $V_{sd}$ and $V_{sq}$ are calculated. The complete FOC system is shown in Fig. 3.9.

### 3.4 Direct Torque Control

DTC method involves no co-ordinate transforms which allows direct control of torque and flux and selecting an optimal switching state. DTC technique provides fast torque response. DTC technique is advantageous over the FOC Technique by the following points

1. No vector transformation
2. Use of open loop flux and torque estimators to improve the performance at low speeds

It is required to maintain short sampling period time in order to maintain the electromagnetic torque ripple within an acceptable hysteresis band.
Fig. 3.9: FOC System for Induction Motor using Matrix Converter [67]
Hence DTC algorithm is usually implemented on a DSP board. DTC allows the compensation of instantaneous errors in flux magnitude and torque under the constraint of unity input power factor. The conventional DTC scheme for selecting the switch state for MC is shown in Fig. 3.10. The command stator flux $\varphi_s^*$ and torque $T_e^*$ magnitudes are compared with the respective estimated values; the errors are then processed through hysteresis band controllers as shown in Fig. 3.10. Using the measured output currents and voltages, the motor flux is estimated and then the electromagnetic torque is estimated.

$$T_e = \frac{3}{2} p (i_{qs} \varphi_{ds} - i_{ds} \varphi_{qs})$$

The estimated electromagnetic torque $T_e$ is compared with electromagnetic torque reference $T_e^*$ using the three level hysteresis comparator and hence decrease, increase or maintain the torque depending on the comparator output. The torque control loop has three digital outputs [69]

$$H_{Te} = 1 \quad for \quad E_{Te} > +HB_T$$
$$H_{Te} = -1 \quad for \quad E_{Te} < -HB_T$$
$$H_{Te} = 0 \quad for \quad -HB_T < E_{Te} < +HB_T$$

where $H_{Te}$ is the output of hysteresis torque controller, $E_{Te}$ is the error between the command torque $T_e^*$ and estimated torque value $T_e$, and $HB_T$ is the total hysteresis band width of the torque controller. In the same way, the actual stator flux can be calculated from the voltage and current information in stationary reference frame as

$$\varphi_{qs} = \int (v_{qs} - i_{qs} r_s) \, dt$$
$$\varphi_{ds} = \int (v_{ds} - i_{qs} r_s) \, dt$$
$$|\varphi_s| = \sqrt{\varphi_{qs}^2 + \varphi_{ds}^2}$$
The estimated flux value ($\phi_e$) is then compared with a command stator flux ($\phi^*_s$) using a two-level hysteresis comparator to increase or decrease the output flux level. It is important to maintain a narrow hysteresis band of the hysteresis comparators to get smooth flux waveforms.

$$H_{\phi} = 1 \quad \text{for} \quad E_{\phi e} > HB_{\phi}$$  \hspace{1cm} (3.67)

$$H_{\phi} = -1 \quad \text{for} \quad E_{\phi e} < -HB_{\phi}$$  \hspace{1cm} (3.68)

where $2HB_{\phi}$ is the total hysteresis-band width of the flux controller, $H_{\phi}$ is the output of the flux controller and $E_{\phi e}$ is the error between the estimated flux $\phi_e$ value and stator flux reference $\phi^*_s$ value. The angular sector of the actual flux is to be calculated to select the proper switching state. The comparator outputs and switching sector are then used to select the voltage vector.

**Fig.3.10: Conventional DTC System[70]**

The selection of switching voltage vector from the flux and torque error values is given in the Table 3.1.
Table 3.1: Selection of Switching Vector in DTC System

<table>
<thead>
<tr>
<th>Sector of Flux</th>
<th>$E_{qe} = 0$</th>
<th>$E_{qe} = 1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$E_{Te} = -1$</td>
<td>$V_2$</td>
<td>$V_3$</td>
</tr>
<tr>
<td>$E_{Te} = 0$</td>
<td>$V_7$</td>
<td>$V_6$</td>
</tr>
<tr>
<td>$E_{Te} = 1$</td>
<td>$V_6$</td>
<td>$V_5$</td>
</tr>
</tbody>
</table>

3.5 Simulation Results and Discussion

3.5.1 FOC System

Fig.3.11 shows the FOC system diagram. The system uses discrete PI controllers. The rotor flux linkage $|\varphi_{r,ref}|$ is maintained at 0.91Wb which is calculated from the rated conditions. The rated reference speed is maintained at 1000rpm. At 400V supply voltage, the maximum stator line-line voltage in MC Drive is 345V. The starting speed of the IM is set at 800 rpm at the input-output voltage transfer ratio of 0.85. The system shown is simulated using Matlab/Simulink. The speed, torque and stator current harmonic spectrum are presented in Fig.3.12, Fig.3.13 and Fig.3.14 respectively. When the reference speed is maintained at 800 rpm, the motor speed increases linearly and reaches the reference value of 800 rpm at 0.9 sec. The torque curve increases to 790 Nm and then settles at 300 Nm until 0.9 sec. The load torque increases to +500 Nm at 1sec. and it is maintained at +500 Nm until 1.5 sec. At 1.5 sec. the speed transition is initiated and the speed changes from 800 rpm to 0 rpm. The speed curve reaches 0 rpm at 2.3 sec. During the speed transition the torque curve maintains +300 Nm from 1.5 sec. to 2 sec. and then decreases to -800 Nm at 2sec. This is due to high fluctuation in the speed curve. At 2.4 sec. the speed curve settles at 0 rpm and the torque curve settles at -500 rpm. The simulation time is 3 sec. It can be
seen from the results that the speed fluctuations are very high. Also it can be seen that the THD level present in the stator current is very high.
Fig. 3.12: Motor Speed Changes

Fig. 3.13: Rotor Torque Changes
3.5.2 Direct Torque Control System

The DTC system is designed using Matlab / Simulink using discrete PI flux and torque controllers as shown in Fig.3.15. The speed variation of IM is shown in Fig.3.16. The simulated cases are motor speed, torque and stator current and their harmonic spectra. The motor speed reaches the reference speed of 800 rpm at 0.9 sec. and it is maintained at 800 rpm until 1.5 sec. From Fig.3.17, it can be seen that the load torque increases to 300 Nm and it is maintained at this value until 0.9 sec. During the constant speed operation from 0.9 sec. to 1.5 sec., the torque decreases to 0 Nm and increases to 500 Nm and it is maintained at +500 Nm until 1.5 sec. The speed transition is initiated at 1.5 sec. and reaches 0 rpm at 2.5 sec. During this time the torque value decreases to 200 Nm and stays at 200 Nm until 2sec. After 2 sec., it decreases to a negative maximum value of 800 Nm. When the speed settles at 0 rpm the torque settles at -500 rpm. The simulation time is 3 sec. It can be seen from the results that torque fluctuations are very high during speed transitions. The initial
torque of the system reaches to 300 Nm. The speed fluctuations are less compared to FOC system. Also it can be seen from Fig.3.18 that the harmonics present in the stator current is very high.
Fig. 3.16: Motor Speed Curve

Fig. 3.17: Motor Torque curve
3.6 Conclusion

The basic modelling and simulation of IM using state space equations ([51] & [95]) is presented in this chapter. The parameters of IM are introduced. The IM model serves as motor parameter observer in FOC and DTC systems.

The basics of modelling and simulation of FOC System and DTC System are presented in this chapter. In addition, the modelling of IM in d-q axis is also presented.

The modelling of FOC system includes the design of speed controller, current controller and flux controller and the complete FOC system is simulated in Simulink. The modelling of DTC System includes design of Torque and Hysteresis controller using PI controllers.
The MC is used as IM drive in FOC and DTC systems. The simulation results of both the control systems show that the presence of PI controllers produces fluctuations and overshoots in speed. The harmonic content present in the stator current is also high.

The implementation of MC in FOC and DTC systems shows that the conventional system using PI controllers leads to speed overshoot and speed fluctuations during torque changes and high number of harmonics present in stator current in FOC and DTC systems. This high number of harmonics and speed and torque fluctuations reduces the efficiency of MC drive for IM.