Chapter 4
Direct Torque Control and Sensor-Less Control of Induction Machine

4.1 Introduction

The application of vector control method for variable speed induction motor drives has been described in chapter 3. Generally, a closed loop vector control scheme results in a complex control structure as it consists of the following components;

1. PID controller for motor flux and toque
2. Current and/or voltage decoupling network
3. Complex coordinate transformation
4. Two axis to three axis transformation
5. voltage or current modulator
6. Flux and torque estimator
7. PID speed controller

In a direct torque control system introduced by Takahashi and Depenbrock independently in 1986, the first five components are replaced by two hysteresis comparators and a selection table[1−3]. This method therefore, results in highly simplified control structure compared to vector control. In the vector control scheme, it is assumed that the controllable power source can force any desired wave shape and value of current into the stator winding. But in practical circuits an inverter can produce only seven discrete space vector values of the actuating variable. In most cases none of these values is exactly equal to the desired instantaneous value of the space vector. Although by using high switching frequency in a PWM inverter the desired curves of the actuating variable can be sufficiently approximated. However, for high power drives, the switching frequency can not be more than 200-300 Hz due to economic reasons. Thus, in high power drives, it is very difficult to apply a current wave

M. Ahmad: *High Performance AC Drives*, Power Systems, pp. 77–96.
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shape of desired magnitude and shape. The vector control therefore, can not provide very fast control required in many drives.

In direct torque control instantaneous values of torque and flux are calculated from primary variables and controlled independently by using an optimum switching table. The controllers for a direct torque control drive do not require complex coordinate transformation essential in all vector controlled drives. Instead the decoupling of non-linear ac motor structure is obtained by the use of 'on –off' control of inverter switches. The voltage vector is selected from the inverter feeding the motor with the help of hysteresis controllers.

4.1.1 Sensorless Control

The vector controlled ac motor drives require speed or position sensors. The presence of sensors results in many disadvantages in terms of cost, reliability, motor size, and noise immunity. The sensorless drives have therefore been developed for precise control of torque and speed. In these drives the speed of the motor is estimated from the applied voltage, line current and frequency. The direct torque control and sensorless drives are very popular now and is a subject of discussion in this chapter. Sensorless drives are now well established in those industrial applications where persistent operation at lower speed is not required.

4.2 Direct Torque Control Basics

The direct torque control is different from vector control in such a way, that it does not control the flux through the current control, but it directly controls the flux itself. The direct torque control is also different from vector control in the sense that the reference frame here is stator flux instead of rotor flux used in vector control.

Direct torque control is in fact an extension of direct vector control and direct self control. Direct vector control theory is quite well known and has been used in industrial drives. As in case of Vector control, in direct torque control method also, the flux and the torque are either measured or estimated and used as feedback signals for the controller. The input to the direct toque controller are the torque error, error in magnitude of the stator flux space vector, and the angle of the stator flux space vector, from which the states of the power switches are determined. Based on this information, a certain voltage vector or combination of voltage vectors is directly applied to the inverter with a certain average timing. This gives the induction motor drives a very fast response.

4.2.1 Torque and Flux Control

The torque in an induction motor in stator reference frame can be expressed as-

$$T_e = \frac{3}{2}\frac{P}{2}\lambda_s^s i_s^s \tag{4.1}$$

In order to apply flux control the stator current i_s^s is to be replaced by rotor flux vector from following expressions.

$$\lambda_s^s = L_s i_s + L_m i_r \tag{4.2}$$

$$\lambda_r = L_r i_r + L_m i_s \tag{4.3}$$

Also

$$\lambda_s = \frac{L_m}{L_r}\lambda_r + (L_s L_r - L_m^2) i_s \tag{4.4}$$

From these equations the torque can be written as-

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r(L_s L_r - L_m^2)}\lambda_r \lambda_s \tag{4.5}$$

The magnitude of torque is

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r(L_s L_r - L_m^2)}\,|\lambda_r|\,|\lambda_s|\sin\gamma \tag{4.6}$$

Where γ is the angle between the stator and rotor flux, known as torque angle. In direct torque control the stator flux is a state variable that is controlled by stator voltage using the following equation

$$\frac{d\lambda_s}{dt} = v_s - R_s i_s \tag{4.7}$$

Here v_s is the inverter output voltage applied to the motor. If the motor is supplied through a voltage source inverter, the inverter voltage vectors with six active vectors, and two zero vectors are shown in Figure 4.1. From equation 4.7, neglecting stator resistance the stator flux is-

$$\lambda_s = \int v_s dt \tag{4.8}$$

For six step operation, the inverter output voltage consists of a cyclic and symmetric sequence of active vectors, so that in accordance with Eq. 4.8 stator flux moves with constant speed along a hexagonal path. The application of zero vectors stops the flux, but does not change its path.

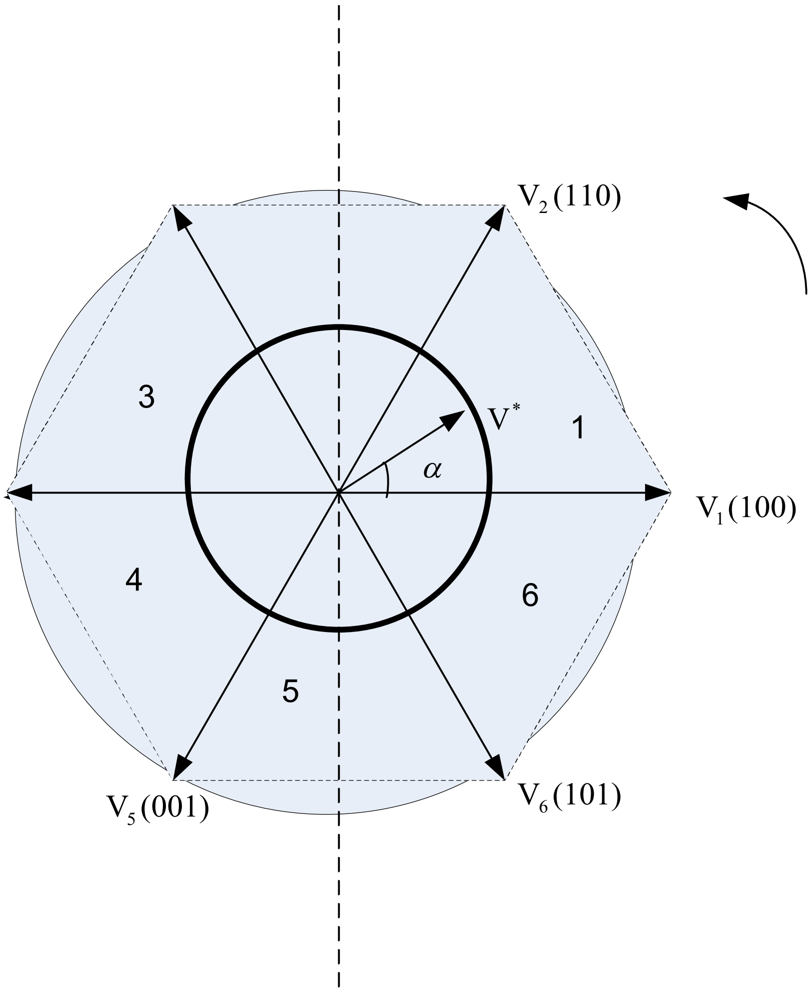

Fig. 4.1 Six active states and two zero states of three-phase inverter

For production of torque the Eq. 4.6 is important, which clearly indicates that the relative angle between the rotor flux and stator flux affects the torque. Suppose the rotor flux λ_r is moving slowly in the anticlockwise direction, if stator flux is moved in clockwise direction by the active voltage vector the angle γ increases rapidly and torque is increased. On the other hand if zero voltage vector is used to stop the stator flux the torque angle and the torque, both will decrease. Thus by cyclic switching of active and zero vectors the torque of the motor can be easily controlled.

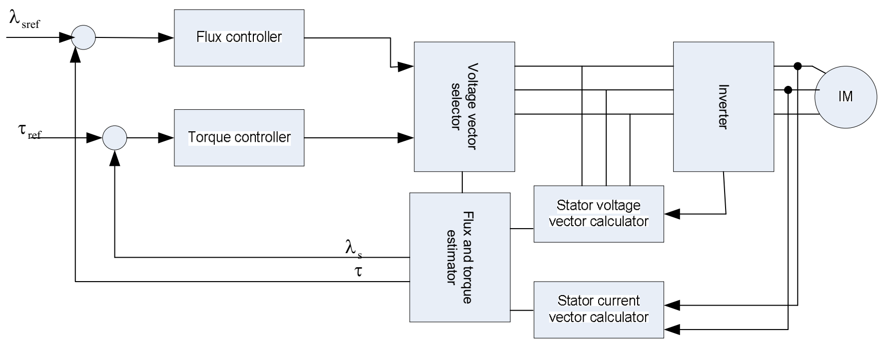

Fig. 4.2 Block diagram of DTC scheme

4.3 DTC Control Strategy

The basic concept of direct torque control as enumerated by Depenbrock and Takahashi is summarized below. The speed of the induction machine mainly depends on the angular speed of the rotating magnetic field. In steady state this speed depends on the number of poles and the frequency of the supply connected to stator. The magnitude of the magnetic flux depends on the voltage to frequency ratio. As shown in Eq. 4.6, if the rotor flux remains constant and stator flux is changed incrementally by the stator voltage, there is corresponding change in angle γ and an incremental change in the torque.

Figure 4.2 shows the basic blocks of a direct torque control scheme. There are three blocks that process the information that is applied to the inverter supplying power to the motor. These blocks are:-

(i) Torque and flux processor
(ii) Optimal switching logic block, and
(iii) Adaptive motor model block.

The torque and flux processing is performed in a hysteresis block. In the torque and flux processor block the reference torque is compared with the actual torque, and the reference flux with the actual flux. The actual values of these quantities are obtained from the adaptive motor model. When the actual torque value drops below its differential hysteresis limit, the torque status output goes high. Similarly, if the actual torque value rises above the differential hysteresis limit, the torque status output goes low. The same function is performed by the flux comparator. The upper and lower differential limit switching points for both torque and flux are determined by the hysteresis window input. This input is used to vary the differential hysteresis limit windows, such that the switching frequencies of the power output devices are maintained within the range of 1.5 to 3.5 KHz.

The voltage vector selector block is an ASIC block that processes the torque status output and flux status output. The function of the optimum switching logic is to select the appropriate stator voltage vector that will satisfy both the torque status output and flux status output. As described in chapter 3, there are eight switching states of a three-phase inverter, out of which two are zero states. The optimal switching table of the inverter is obtained using the torque and flux comparator outputs, and the position of the stator flux linkage space vector.

For controlling the torque of the motor, the inverter switching states are selected based on the following criterion. The torque developed is proportional to the cross product of stator flux vector λ_s and the rotor flux vector λ_r. The stator flux vector is kept constant, and the torque is controlled by varying the angle between the stator flux vector and rotor flux vector. Since the time constant of rotor is much higher than the stator time constant, the rotor flux does not change much during the time of interest. If an increase in motor torque is required due to increase in load torque, the optimal switching logic selects a stator (inverter) voltage vector such that the angle between stator flux vector and rotor flux vector increases. When a decrease in torque is required, the optimum switching logic selects a zero voltage vector, which allows both stator flux and produced torque to decay naturally. If stator flux decays below its normal lower limit, the flux status output requests an increase in stator flux. If the torque status output is low, anew stator voltage vector is selected that tends to increase the flux but reduces the angle γ between the stator and rotor flux vectors.

The combination of torque and flux comparators, combined with optimal switching logic eliminates the need for traditional PWM modulator. This presents major advantage, as the small signal delay associated with modulator are eliminated.

The Adaptive motor model is required to calculate the actual values of stator flux, torque produced, reference speed, and frequency. The actual flux and torque values are critical to direct torque control. These values are therefore calculated every 25 microsecond. The reference speed and frequency are required by the outer speed loop and are calculated every millisecond. The motor adaptive model is developed from the motor parameters and from the measured stator currents, link voltage, and power switches positions The DSP is responsible for calculating all motor variables required in DTC.

Determination of actual stator flux is the most important step in the adaptive motor model. the following two equations are used to determine the stator flux.

$$\lambda_s = \int (v_s - R_s i_s)dt \tag{4.9}$$

and

$$\lambda_s = L_s i_s + L_m i_r \tag{4.10}$$

The initial estimate of the flux is obtained from first equation which is then fine tuned using second equation. Once the flux is determined the torque is calculated using the following equation

$$T_e = K\lambda_s \times i_s \tag{4.11}$$

In order to obtain the values of actual frequency and speed, the rotor flux vector and rotor flux angle is calculated using the following equations-

$$\lambda_r = \frac{L_r}{L_m}(\lambda_s - \sigma L_s i_s) \quad = \lambda_{rd} + j\lambda_{rq} \tag{4.12}$$

where σ is the leakage factor.

Rotor flux angle

$$\theta_r = \arctan\frac{\lambda_{rq}}{\lambda_{rd}} \tag{4.13}$$

The complete block diagram of direct torque control drive is shown in Figure 4.2. The speed control block compares the reference speed with the actual speed feedback from the motor adaptive model. This block contains the traditional PID controller. The output of this block is the reference torque (speed) for the torque controller. The torque controller has one more input as absolute torque reference. If the drive is used for speed control the speed (torque) reference is used. If torque control drive is used then only the absolute torque reference is used. The reference torque is compared with the actual torque in the hysteresis torque controller as discussed earlier. The flux level reference and actual frequency feedback are the inputs to flux reference control block. The actual frequency feedback is used to provide frequency sensitive flux manipulation. The flux reference is compared with the actual flux in flux hysteresis comparator as shown in Figure 4.2. The block showing switching frequency control is used to limit the switching frequency within a minimum and maximum level. The minimum frequency is decided on the basis of voltage waveform, and the maximum frequency is decided based on the switching frequency of the power devices.

The function of the switching frequency control is to vary the size of hysteresis windows of the flux and torque comparators to limit the switching frequencies within 1.5 to 3.5 K Hz. Compared to Vector control (FOC) scheme, the DTC scheme has the following features.

- There are no current loops; hence the current is not regulated directly.
- Coordinate transformation is not required.
- There is no separate voltage pulse width modulation
- Stator flux vector and torque estimation is required.

Depending on how the switching sectors are selected, two different DTC switching schemes are used. The scheme proposed by Takahashi operates with circular stator flux vector path, whereas the scheme of Depenbrock operates with hexagonal stator flux vector path. The two switching sector selections are shown in Figure 4.3.

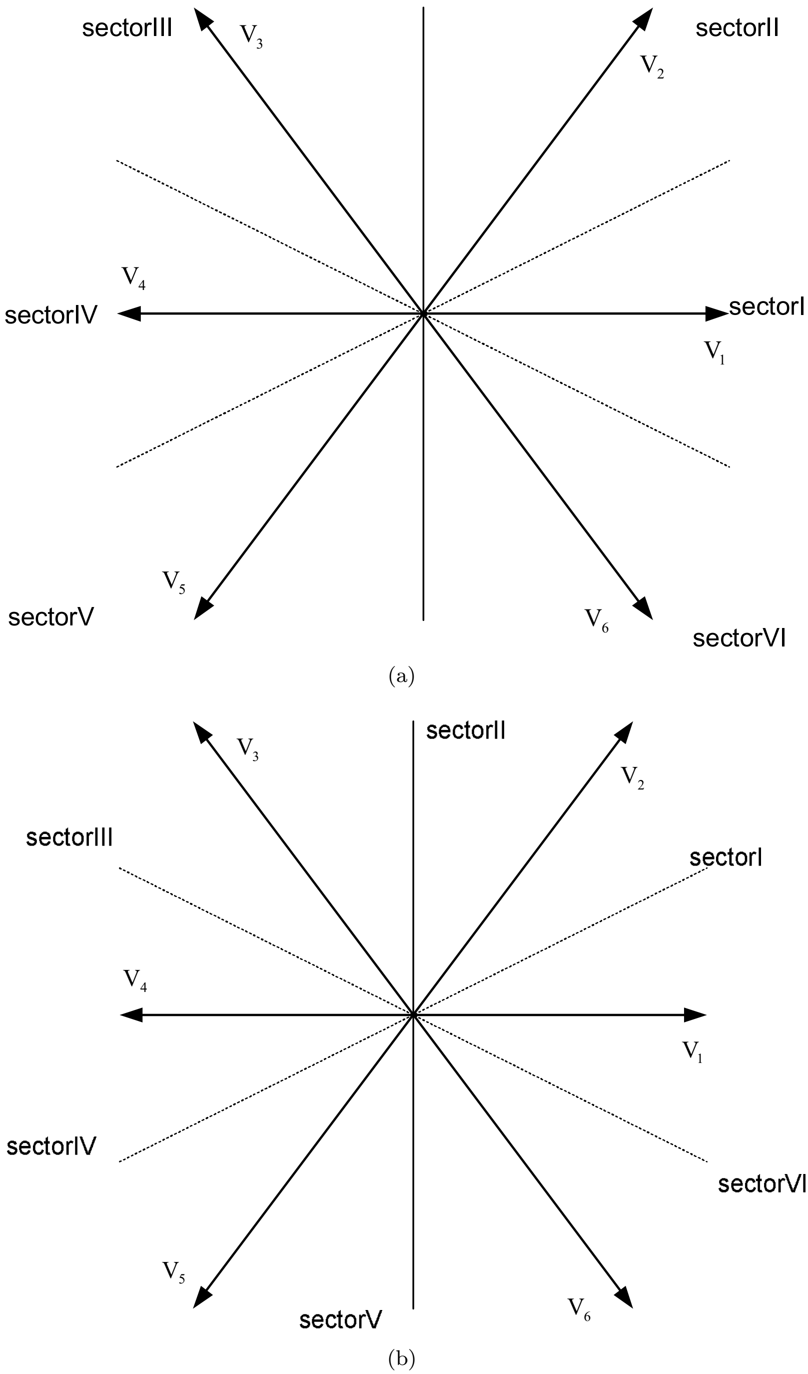

Fig. 4.3 (a). Circular stator flux vector path **(b).** Hexagonal stator flux vector path

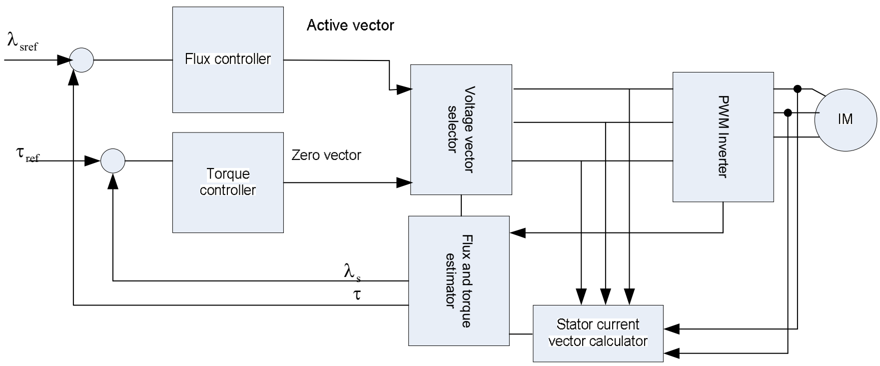

Fig. 4.4 Switching Table based DTC scheme

4.4 Switching Table Based DTC Scheme

The Switching table based direct torque control scheme with circular stator flux vector path is shown in Figure 4.4. The command stator flux and torque values are compared with actual flux and torque values in hysteresis flux and torque controllers respectively. The flux controller is a two- level comparator while the torque controller is a three-level comparator. If the output of torque hysteresis comparator is denoted by τ, then τ= -1 means that the actual value of the torque is above the reference value and outside the hysteresis limit. τ=1 means that the actual value of torque is less than the reference value and outside the hysteresis limit. τ=0 means the torque is within the hysteresis limit.

The flux hysteresis comparator output is denoted by ϕ. If ϕ=0, it means that the actual value of the flux linkages is above the reference value and outside the hysteresis limit. ϕ=1 means the actual value of the flux linkage is below the reference value and outside the hysteresis limit. These digitized variables representing the output of torque and the stator flux region R obtained from the angular position $\gamma = \text{Arctan}(\frac{\lambda_{sq}}{\lambda_{sd}})$ create a digital word which is used as address for accessing EPROM. The EPROM contains the selection table as shown in Table 4.1.

Table 4.1

		R(1)	R(2)	R(3)	R(4)	R(5)	R(6)
$\phi = 1$	$\tau = 1$	$V_2(110)$	$V_3(010)$	$V_4(011)$	$V_5(001)$	$V_6(101)$	$V_1(100)$
	$\tau = 0$	V(000)	V(000)	V(000)	V(000)	V(000)	V(000)
	$\tau = -1$	$V_6(101)$	V_1	V_2	V_3	V_4	V_5
$\phi = 0$	$\tau = 1$	$V_3(110)$	$V_3(010)$	$V_4(011)$	$V_5(001)$	$V_6(101)$	$V_1(100)$
	$\tau = 0$	V(000)	V(000)	V(000)	V(000)	V(000)	V(000)
	$\tau = -1$	V_5	V_6	V_1	V_2	V_3	V_4

The characteristic features of this scheme are summarized below.

- The stator flux and current waveforms are nearly sinusoidal; the harmonic content depends on the flux and torque controller hysteresis bands.
- Excellent dynamic torque behavior.
- The inverter switching frequency is determined by the flux and torque hysteresis bands.

Number of modifications has been made on the basic ST-DTC scheme to improve its performance by using modified switching table. Torque ripples can be reduced by dividing the sampling periods into two or three equal intervals thereby increasing the switching voltage vectors to 12 or 56. Also rotor flux amplitude may be controlled to increase the overload torque capability.

4.4.1 Direct Self Control Scheme

The block diagram of DSC scheme proposed by Depenbrock is shown in Figure 4.5. The flux and torque estimator block has voltage and currents as input. The output of this block is the flux and torque. From the flux values thus obtained, the flux comparators generate digitized variables d_A, d_B *and* d_C which correspond to active voltage vectors for six-step operation of the inverter. The hysteresis torque controller generates the digitized value d_0 that

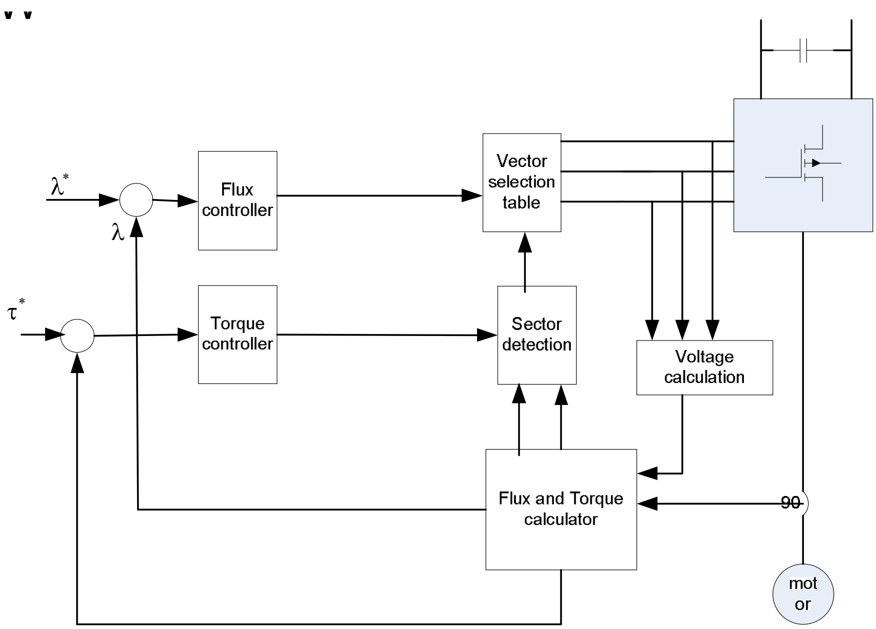

Fig. 4.5 Depenbrock scheme for DTC

determines the zero state duration. The control algorithm for constant flux region is therefore of the form-

If $d_0 = 1$, then inverter switching states are
$S_A = d_B, S_B = d_C$ and $S_C = d_A$ that means active vector is selected.
For $d_0 = 0$ zero vector is selected, ie.

$$S_A = 0, S_B = 0, S_C = 0 \text{ or } S_A = 1, S_B = 1, S_C = 1$$

The main features of DSC scheme are summarized below.

- The DSC scheme has a PWM operation in constant flux region and six step operation in flux weakening region.
- Non-sinusoidal current and flux waveforms in both the regions.
- Stator flux vector moves in hexagonal path in PWM operation also.
- The inverter frequency is lower than in Switching Table based DTC.
- However the behavior of DSC based DTC can be obtained from Switching Table based DTC scheme by properly selecting the hysteresis band of stator flux comparator.

4.4.2 Main Features of DTC

The main features of DTC are summarized below-

- DTC operates with closed torque and flux loops, but without current controllers.
- DTC needs stator flux and torque estimation. It is therefore insensitive to rotor parameters.
- DTC is basically a motion sensorless scheme.
- DTC has simple and robust control structure.

4.5 Sensorless Control of Induction Motor

Vector control schemes of induction motor require speed or position sensor. The speed sensor has several disadvantages in terms of cost, reliability, drive size, and noise immunity. As real time computation costs are continuously decreasing, speed and position estimation can be performed by using software based state estimation techniques. Various approaches have been proposed in the literature for estimation of speed using stator voltage, phase currents and frequency measurements. Techniques for obtaining the speed information of an induction motor without using the speed encoder can be broadly classified as-

1. Open loop speed control with slip compensation.
2. Closed loop control with speed estimation

In the first case motor synchronous speed is regulated and the estimated slip frequency is used to compensate for load changes. In second method, the motor speed is estimated and used as a feedback signal for closed loop speed regulation. Based on the methods of implementation following techniques can be used for sensorless control of induction motor drives employing vector/ direct torque control.

1. Slip frequency calculation method
2. Speed estimation using state equations
3. Flux estimation method
4. Model reference adaptive systems (MRAS)
5. Observer (Kalman, Luenberger) based methods
6. Artificial intelligence methods for speed estimation

Slip Frequency Calculation Methods

The slip frequency of an induction motor is the difference between the stator frequency and the electrical frequency corresponding to rotor speed. By calculating the slip frequency, the speed of the motor can be obtained. The slip frequency ω_{sl} is related to stator frequency as

$$\omega_{sl} = \omega_e - \omega_r \tag{4.14}$$

Where, ω_e is the stator frequency, and ω_r is the frequency of rotor. A number of methods have been suggested for calculation of slip frequency. In one method[5] the rotor frequency is calculated directly from the phase lag between the stator voltage and the stator current. The slip frequency is then calculated using motor parameters and stator current. From the steady state equations obtained from the equivalent circuit of the induction motor the following relationship between the rotor frequency ω_r and the phase angle ϕ is obtained.

$$\phi = \arctan \frac{D\omega_r^2 - E}{(A\omega_r^2 + B\omega_r + C)} \tag{4.15}$$

Where $A=R_1L_2^2$; $B=\omega_1L_m^2R_2$; $C=R_1R_2^2$; $D=\omega_1(L_2L_m^2 - L_1L_2^2)$; and $E=\omega_1L_1R_2^2$.

From above equation a curve between slip frequency and phase angle can be obtained. The Eq. (4.15) is valid for limited range of slip frequency and can be used when fast dynamic response is not required.

Speed Estimation Using State Equations

The slip frequency is obtained as a function of rotor EMF obtained in stationary reference frame. In this method only measurement of phase voltages and currents is required[6,7] The state equations are modified to express speed in terms of motor parameters and measured quantities.

The stator and rotor voltage equations of the induction machine in the stationary reference frame can be written as-

$$\mathrm{v_{ds}} = \mathrm{R_s i_{ds}} + \frac{\mathrm{d}}{\mathrm{dt}}\lambda_{\mathrm{d}s} \tag{4.16}$$

$$\mathrm{v_{qs}} = \mathrm{R_s i_{qs}} + \frac{\mathrm{d}}{\mathrm{dt}}\lambda_{\mathrm{q}s} \tag{4.17}$$

$$0 = \mathrm{R_r i_{dr}} + \frac{d\lambda_{dr}}{dt} + \omega_r\lambda_{qr} \tag{4.18}$$

$$0 = \mathrm{R_r i_{qr}} + \frac{d\lambda_{qr}}{dt} - \omega_r\lambda_{dr} \tag{4.19}$$

$$\mathrm{T_e} = \frac{3}{2}\frac{\mathrm{P}}{2}(\mathrm{i_{qs}}\lambda_{\mathrm{ds}} - \mathrm{i_{ds}}\lambda_{\mathrm{qs}}) \tag{4.20}$$

and

$$\frac{d\omega_r}{dt} = \frac{1}{M}(T_e - T_L) \tag{4.21}$$

From the above dynamic equations, the expression for slip frequency can be obtained as-

$$\omega_{sl} = \omega_e.R_r\left(\frac{L_m}{L_r}\right)\frac{e_{qr}.i_{qs} + e_{dr}.i_{ds}}{e_{dr}^2 + e_{qr}^2} \tag{4.22}$$

Where the rotor EMFs can be expressed in terms of inductances as-

$$\mathrm{e_{dr}} = \frac{\mathrm{d}\lambda_{\mathrm{dr}}}{\mathrm{dt}} = \left(\frac{\mathrm{L_m}}{\mathrm{L_r}}\mathrm{R_r i_{ds}} - \frac{\mathrm{R_r}}{\mathrm{L_r}}\lambda_{\mathrm{dr}} - \omega_{\mathrm{r}}\lambda_{\mathrm{qr}}\right) \tag{4.23}$$

$$\mathrm{e_{qr}} = \frac{\mathrm{d}\lambda_{\mathrm{qr}}}{\mathrm{dt}} = \left(\frac{\mathrm{L_m}}{\mathrm{L_r}}\mathrm{R_r i_{qs}} - \frac{\mathrm{R_r}}{\mathrm{L_r}}\lambda_{\mathrm{qr}} + \omega_{\mathrm{r}}\lambda_{\mathrm{dr}}\right) \tag{4.24}$$

Model Reference Adaptive System

Model reference Adaptive Systems approach makes a comparison of two machine models of different structure, that estimate the same state variable on the basis of different set of input variables. One model does not include speed and is called the reference model. The other, which includes the speed also, is called the adjustable model. The error between the two models is used to derive an adoption model that produces the estimated speed for the adjustable model. The adjustments in the adjustable models are made so that the error between the two models vanishes to zero. A block diagram for estimation of speed using MRAS technique is shown in Fig. 4.6. The reference model is obtained as stator voltage equations. The inputs to this model are motor stator voltage and current signals. The output is calculated in the form of rotor

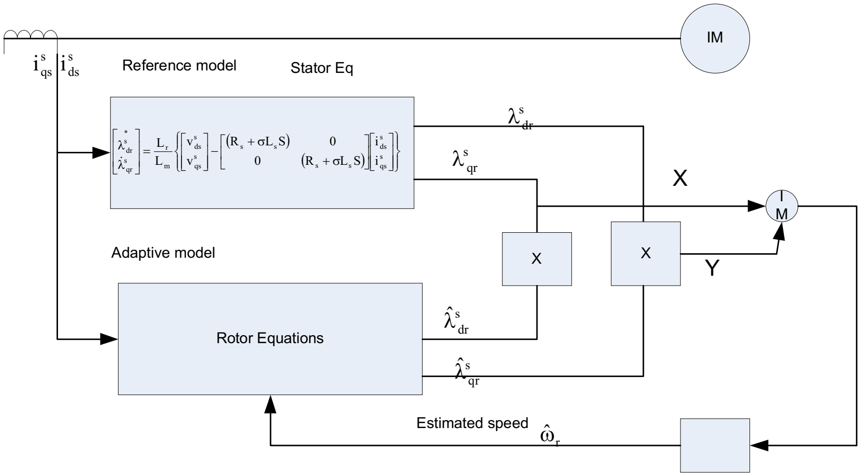

Fig. 4.6 Speed estimation for MRAS system

flux vector λ^s_{dr} and λ^s_{qr}. The adaptive model of the motor is obtained using stator current inputs and an estimated speed signal (assuming it is available). The output from this model is also the rotor flux linkages $\tilde{\lambda}^s_{dr}$ and $\tilde{\lambda}^s_{qr}$. If the estimated speed is correct the fluxes calculated from the reference model and the adaptive model must have the same value. An adaptation algorithm with PI control can be used to tune the speed $\tilde{\omega}_r$ so that the error $\xi = 0$. The PI control is implemented as-

$$\tilde{\omega}_r = \xi(K_p + \frac{K_I}{s}) \tag{4.25}$$

and

$$\xi = (\lambda^s_{qr}\tilde{\lambda}^s_{dr} - \lambda^s_{dr}\tilde{\lambda}^s_{qr}) \tag{4.26}$$

Luenberger Speed Observer

A device that estimates or observes state variables of a system is called a **state observer**. A state observer utilizes measurements of the system inputs and outputs and a model of the system based on differential or difference equations. Three main quantitative state observers are: Luenberger observer, adaptive observer and Kalman filter. In the deterministic case, when no random noise is present, the Luenberger observer and its extensions are used for time-invariant systems with known parameters. The equation for the Luenberger observer contains a term that corrects the current state estimates by an amount proportional to the prediction error: the estimation of the current output minus the actual measurement. An observer is basically an estimator that uses full or partial plant model, and a feedback loop with measured plant variables. In the case of speed observer first the stator currents i^s_{ds} and i^s_{qs} and the rotor fluxes λ^s_{dr} and λ^s_{qr} are calculated through a full order Luenberger observer based on stator and rotor equations in stator coordinates. The rotor voltage equations can be written from voltage model as-

$$v^s_{dr} = 0 = i^s_{dr}R_r + \frac{d}{dt}(\lambda^s_{dr}) + \omega_r\lambda^s_{qr} \tag{4.27}$$

$$v^s_{qr} = 0 = i^s_{qr}R_r + \frac{d}{dt}(\lambda^s_{qr}) + \omega_r\lambda^s_{dr} \tag{4.28}$$

Also

$$\lambda^s_{dr} = L_m i^s_{ds} + L_r i^s_{dr} \tag{4.29}$$

$$\lambda^s_{qr} = L_m i^s_{qs} + L_r i^s_{qr} \tag{4.30}$$

By eliminating i^s_{dr} the following equation is obtained.

$$\frac{d}{dt}\lambda^s_{dr} = -\frac{R_r}{L_r}\lambda^s_{dr} - \omega_r\lambda^s_{qr} + \frac{L_m R_r}{L_r} i^s_{ds} \tag{4.31}$$

Similarly for q axis

$$\frac{d}{dt}\lambda^s_{qr} = -\frac{R_r}{L_r}\lambda^s_{qr} + \omega_r\lambda^s_{dr} + \frac{L_m R_r}{L_r} i^s_{qs} \tag{4.32}$$

The stator currents i^s_{ds} and i^s_{qs}, in terms of machine parameters can be expressed as

$$\frac{d}{dt}(i^s_{ids}) = -\left(\frac{L_m^2 R_r + L_r^2 R_s}{\sigma L_s L_r^2}\right) i^s_{ds} + \frac{L_m R_r}{\sigma L_s L_r^2}\lambda^s_{dr} + \frac{L_m\omega_r}{\sigma L_s L_r}\lambda^s_{qr} + \frac{1}{\sigma L_s} v^s_{ds} \tag{4.33}$$

$$\frac{d}{dt}(i^s_{iqs}) = -\left(\frac{L_m^2 R_r + L_r^2 R_s}{\sigma L_s L_r^2}\right) i^s_{qs} - \frac{L_m R_r}{\sigma L_s L_r}\lambda^s_{dr} + \frac{L_m\omega_r}{\sigma L_s L_r^2}\lambda^s_{qr} + \frac{1}{\sigma L_s} v^s_{qs} \tag{4.34}$$

Where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$.

The state variable equations can now be written in the form of

$$\left[\dot{X}\right] = [A]\,[X] + [B]\,[U] \tag{4.35}$$

where

$$[X] = \left[\, i^s_{ds}\ i^s_{qs}\ \lambda^s_{ds}\ \lambda^s_{qs} \,\right]^T \tag{4.36}$$

$$[U] = \left[\, v^s_{ds}\ v^s_{qs}\ 0\ 0 \,\right]^T \tag{4.37}$$

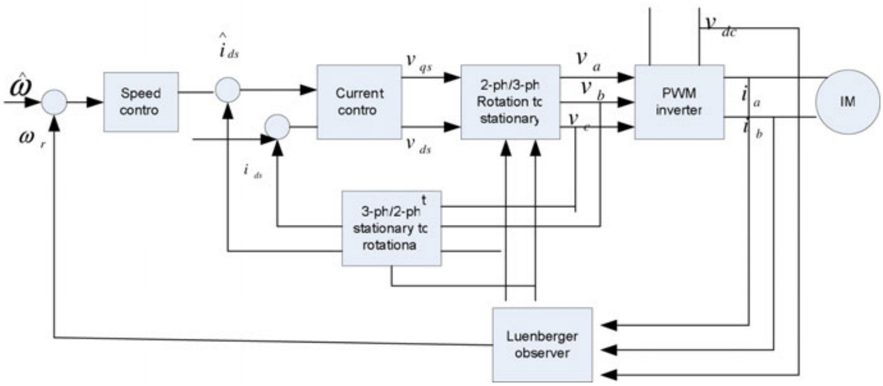

Fig. 4.7 Block Diagram of Luenberger method of speed control

$$[A] = \begin{bmatrix} \frac{-\left(L_m^2 R_r + L_r^2 R_s\right)}{\sigma L_s L_r^2} & 0 & \frac{(L_m R_r)}{\sigma L_s L_r^2} & \frac{L_m \omega_r}{\sigma L_s L_r} \\ 0 & \frac{-\left(L_m^2 R_r + L_r^2 R_s\right)}{\sigma L_s L_r^2} & \frac{-L_m \omega_r}{\sigma L_s L_r} & \frac{L_m R_r}{\sigma L_s L_r^2} \\ \frac{L_m R_r}{L_r} & 0 & -\frac{R_r}{L_r} & -\omega_r \\ 0 & \frac{L_m R_r}{L_r} & \omega_r & \frac{R_r}{L_r} \end{bmatrix} \tag{4.38}$$

$$[B] = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \tag{4.39}$$

Fig. 4.7 shows the block diagram of Luenberger observer using the above machine model. The estimated values are shown with $\wedge$. The output current signals $\hat{i}_{ds}$ and $\hat{i}_{qs}$ are obtained from the following equation (Here superscript 's' is not written)

$$\begin{bmatrix} \hat{i}_{ds} \\ \hat{i}_{qs} \end{bmatrix} = \begin{bmatrix} 1\,0\,0\,0 \\ 0\,1\,0\,0 \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \\ \lambda_{dr} \\ \lambda_{qr} \end{bmatrix} \tag{4.40}$$

The input voltage signals $v_{ds} \mathrm{and} v_{qs}$ are measured from stator terminal of the machine. As shown in Fig. 4.8, the speed adaptation algorithm utilizes the speed adaptive flux observer obtained from the machine model. The observer equation is given by

$$\frac{d}{dt}(\hat{X}) = \hat{A}\hat{X} + BV_s + G(\hat{i}_s - i_s) \tag{4.41}$$

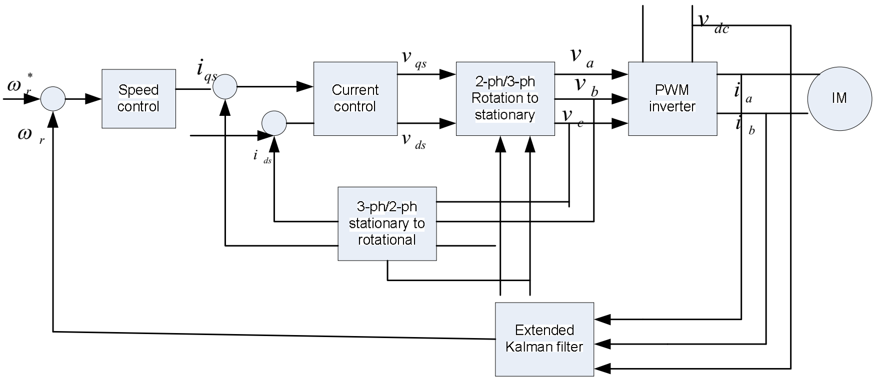

Fig. 4.8 Block Diagram of Extended Kalman Filter method of speed control

where $\hat{i}_s = [i_{ds} i_{qs}]$ and G observer gain matrix. The gain matrix G multiplied with the error signal $e = \hat{i}_s - i_s$ applies the corrective signal such that e becomes zero. If speed signal ω_r in matrix A is known the fluxes and currents can be solved from the state equations. The speed adaptive flux observer permits estimation of the unknown speed ω_r. The estimation error in the stator currents and rotor fluxes is expressed by the equation:

$$\frac{d}{dt}(e) = (A + GC)e - \Delta A\hat{X} \tag{4.42}$$

Where

$$e = X - \hat{X}, \quad \Delta A = \hat{A} - A, \text{ and } \Delta\omega_r = \hat{\omega}_r - \omega_r \tag{4.43}$$

To derive the speed adaptive algorithm the following Lyapunov function is considered:

$$V = e^T e + \frac{(\hat{\omega}_r - \omega_r)^2}{\lambda} \tag{4.44}$$

Using Lyapunov's theorem the following adaptation scheme for speed estimation can be obtained,

$$\frac{d\hat{\omega}_r}{dt} = \lambda(e_{ids}\hat{\lambda}_{qr} - e_{iqs}\hat{\lambda}_{dr})L_m/(\sigma L_s L_r) \tag{4.45}$$

here $e_{ids} = i_{ds} - \hat{i}_{ds}$ and $e_{iqs} = i_{qs} - \hat{i}_{qs}$.

The gain matrix is so chosen as to make the speed adaptive flux observer stable. In this method the estimation error becomes large at low speeds due to the effects of stator and rotor resistance variation.

Extended Kalman Filter Observer

The Luenberger observer is a deterministic observer, and is applicable to linear time-invariant systems In the extended Kalman filter, (EKF) the state transition and observation models need not be linear functions of the state but may instead be (differentiable) functions. In EKF the state variables are only the stator currents and the magnetizing current. The block diagram of EKF algorithm is shown in Fig. 4.8. The induction motor model in stationary reference frame with stator currents i_{ds}, i_{qs} and rotor fluxes $\lambda_{dr}, \lambda_{qr}$ as state variables is as foillows-

$$\frac{dX}{dt} = AX + BU \tag{4.46}$$

and

$$Y = CX \tag{4.47}$$

Where

$$X = \begin{bmatrix} i_{ds} & i_{qs} & \lambda_{dr} & \lambda_{qr} & \omega_r \end{bmatrix}^T \tag{4.48}$$

$$Y = \begin{bmatrix} i_{ds} & i_{qs} \end{bmatrix}^T \tag{4.49}$$

$$[B] = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \tag{4.50}$$

$$[A] = \begin{bmatrix} \frac{-\left(L_m^2 R_r + L_r^2 R_s\right)}{\sigma L_s L_r^2} & 0 & \frac{(L_m R_r)}{\sigma L_s L_r^2} & \frac{L_m \omega_r}{\sigma L_s L_r} & 0 \\ 0 & \frac{-\left(L_m^2 R_r + L_r^2 R_s\right)}{\sigma L_s L_r^2} & -\frac{L_m \omega_r}{\sigma L_s L_r} & \frac{(L_m R_r)}{\sigma L_s L_r^2} & 0 \\ \frac{L_m R_r}{L_r} & 0 & -\frac{R_r}{L_r} & -\omega_r & 0 \\ 0 & \frac{L_m R_r}{L_r} & \omega_r & -\frac{R_r}{L_r} & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix} \tag{4.51}$$

$$C = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 \end{bmatrix} \tag{4.52}$$

In the dynamic model of an induction motor, if the dimension of state vector is increased, by adding the angular speed of motor, then state model is of fifth order, as well it becomes non-linear. In this case the speed of the rotor is considered as a state and a parameter. The extended Kalman filter algorithm is calculated using a microprocessor, and the system is expressed in a discrete form. The discrete model of the induction motor with noise sources is

$$X(k+1) = A_d X(k) + B_d U(k) + V(k) \tag{4.53}$$

$$Yk) = C_d X(k) + W(k) \tag{4.54}$$

Where $W(k)$ *and* $V(k)$ are zero mean white Gaussian noise vectors of Y(k) and X(k) respectively. Both $W(k)$ *and* $V(k)$ are independent of Y(k) and X(k) respectively. The statistics of noise and measurements are given by three covariance matrices, Q, R and P. The system noise vector covariance matrix, and system state vector covariance matrix P are 5×5 matrices, whereas the measurement noise vector covariance matrix is 2×2 only. The EKF has two main stages: the prediction stage and filtering stage. In prediction stage, the next predicted values of states $X^*(k+1)$ are calculated by the machine model and the previous values of estimated states. The prediction of state is given by the following equation

$$\mathrm{X}^*(\mathrm{k}+1) = \mathrm{X}^*(\mathrm{k}) + \int_{\mathrm{t_k}}^{\mathrm{t_{k=1}}} \mathrm{f}[\mathrm{x}(\mathrm{t}), \mathrm{U}(\mathrm{t})\mathrm{dt} \tag{4.55}$$

Here U(t) is assumed to remain constant during t_k *to* t_{k+1}. The predicted state covariance matrix $P^*(k+1)$ is obtained using the system noise covariance vector Q.

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