Chapter 8

Development, Simulation and Application of ANN Based IM Model

Chapter 8:
DevelopmEnt, Simulation and ApPlication of ANN Based Induction MachIINe Model

Chapter covers a modular SIMULINK implementation of an induction machine model, described in a step-by-step approach. With the modular system, each block solves one of the model equations; therefore, unlike black box models, all of the machine parameters are accessible for control and verification purposes. The model is used in different drive applications, such as open-loop constant V/HZ control and indirect vector control are given. The use of the model as an induction generator is demonstrated. The neural network is used as an estimator to estimate the required machine parameters.

8.1 CHARACTERISTICS OF INDUCTION MOTOR

The performance characteristics of a three phase induction motor can be derived using the approximate equivalent circuit shown in fig 8.1

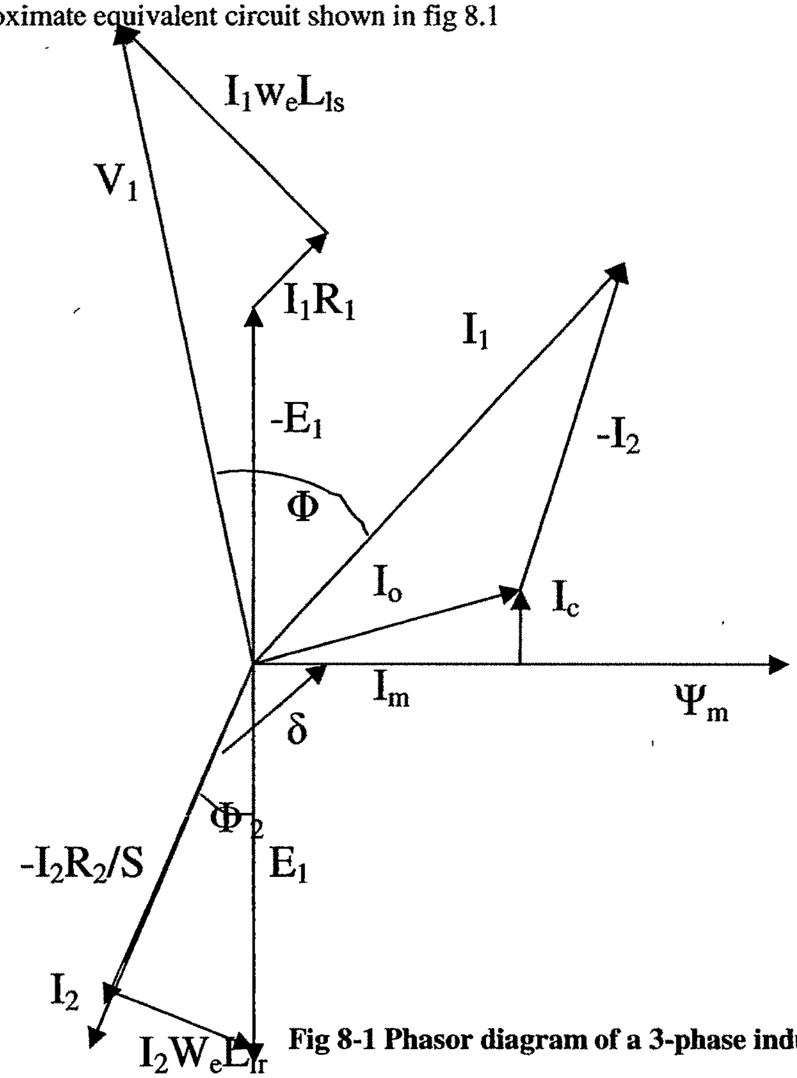


Fig 8-1 Phasor diagram of a 3-phase induction motor

In an induction motor the power transferred to the rotor(Pd), rotor copper Loss (Pcu2) and mechanical power developed (Pm) are in the ratio of 1:s :(1-s).

The torque developed by the motor

$$Td = m1\ Pm/2\prod nr\ \ Nm \tag{8.1}$$

This can be simplified by

$$Td = m1\ v1^2 r2'/(s2\pi n_s(r1 = r2'/s)^2 + (x1+x2')^2) \tag{8.2}$$

At very small slips (operating region of the motor)

$$(r1+r1'/s) >> (x1+x2') \tag{8.3}$$

also: r2'/s>>r1 leading to

$$Td = m1\ (v1^2 s/2\pi nsr2') \tag{8.4}$$

This shows that torque developed is directly proportional to the slip. In a similar way, at large slips the torque varies in inverse proportion to the Slip. It is known that there is a maximum value for the torque. The slip at which this maximum torque occurs is given by:

$$S_m(T) = \pm 2' /\sqrt{}\ (r1^2+(x1+x2')^2) \tag{8.5}$$

The maximum torque is:

$$T_{dm} = m_1\ V_1^2/\ (2\pi n_s)\ 2[r1 \pm \sqrt{}\ (r1^2+(x_1+x_2')\ ^2] \tag{8.6}$$

Sometimes the stator impedance is neglected of the stator impedance drop is compensated to operate the motor at constant flux at all slips. The applied voltage is increased such that the induced voltage E1/f is constant (equal to the value at rated voltage). The applied voltage varies as a function of frequency such the E1/f remains constant. The stator impedance can be assumed as a zero, since it has no effect. In such case the relations are

$$Td = m1E1^2/2\pi ns\ 2x2' \tag{8.7}$$

The characteristics features are:

- At exactly synchronous speed s=0, the torq2ue developed is zero (Td=0). This is expected because no induced currents due to zero relative speed.
- Full load torque (T_{fl}) corresponds to the rated sip (s_{fl}).
- T_{dm} is maximum torque at the slip s_m (T).
- T_{st} is the starting toque at s=1.

$$T_d/t_{dm} = 2/(s\ /\ s_m + s_m\ /\ s) \tag{8.8}$$

The operation of the motor in the range of slips 0-sm is stable. When the motor is operating in this range any disturbance in the operating point by the change of either speed or torque is damped out and the motor returns to its operating point, or attains a new one. For stable operation the torque developed must be increase when the speed falls, that is, DT_d/dn_s should be negative. The operation of the motor in the range sm-1 is unstable. In this range the curve has the positive dT_d/dn_s., i.e. Torque decreases when speed falls. The characteristic is almost linear at very small slips (in the stable operation region). This linearity continues till the breakdown shown in fig is redrawn in the same fig.

For slips greater than unity, the operation is in the fourth quadrant. The rotation of the rotor and the rotating magnetic field are in opposite direction. The torque developed is the breaking torque, tending to stop the motor. This can occur in two ways.

- The phase sequence of the supply to the motor is reversed while it is running.
- A negative torque is applied to the shaft.

The motor is operated as a brake in the range of slip (s>1), to make it drive the load at constant speed, while lowering the load. The torque is positive whereas the direction of the rotation is reversed. By suitable resistance of the rotor the point of operation is shifted to the quadrant of operation, so that the load can be lowered at constant speed.

The torque speed curve extends to the second quadrant, representing the negative torque in the forward direction of rotation. This occurs if the speed of the rotor is greater than the synchronous speed is arrested by the generating torque. In the mode of operation all the kinetic energy connected with increase in speed is returned to the mains. The maximum (breakdown) torque depends on the following.

- It varies as the square of the applied voltage.
- It Decreases with stator impedance.
- Its value is independent to rotor resistance.
- Its value decreases with increase in rotor leakage resistance.

8.2 Desirable modifications

The section describes modifications in the characteristics of IM for various applications

- **Additional rotor resistance.**

Maximum torque is independent of rotor resistance, but the slip at which the maximum torque occurs changes with rotor resistance. Starting torque can be increased by increasing the rotor resistance. But the running characteristics are impaired due to low efficiency impaired by high rotor resistance due to increased losses. Rotor heating is present in an in inherently high resistance motor.

- **Variation Of Applied Voltage**

Typical speed torque characteristic for induction motor when supplied from variable voltage are shown in the figure at rated frequency. They are based on the fact that the induction motor torque (at a given slip) varies as the square of voltage. The slip for maximum torque is independent of voltage. The full load torque occurs at different slips when the voltage is varied. This renders the speed control of induction motor feasible over a limited range by supply voltage variation. Due to reduction in air gap flux at low voltages, the torque capability of the motor decreases. The power factor decreases. The motor draws the heavy currents to develop a given torque at low voltage. The current drawn at different voltages is shown in the fig. Along with the torque developed at rated current at different voltages. Fig (b) shows the advantages of high resistance in the rotor when the applied voltage is varied to modify the speed. Speed torque characteristic. Besides increasing the range of speed control, the current drawn by the motor at low voltages can be limited by proper choice of rotor resistance.

- **Pole Change Motor**

The method is more suitable for squirrel cage motor as their rotor can adapt to any number of poles. The speed changes with the change in number of poles. The type of connection decides the permissible loading at constant torque and power.

- **Slip Power Recovery**

Variable rotor resistance method at continuous low speed operation becomes uneconomical due too overheating of the rotor. These low speeds can be very effectively achieved with reasonable efficiency using slip energy recovery schemes. The slip power which is wasted in the external resistance is returned to the mains in these schemes.

When these methods are employed, the motor may be operated to drive both constant torque and constant power loads. Such schemes are scheribous and Kramer controls. If the slip power converter allows flow in both direction, the motor may be operated at both super and sub synchronous speeds. In super synchronous operation the power at line frequency is converted to slip power and fed to motor. While at sub synchronous speeds the slip power is converted to line frequency and fed to the mains.

- **Injection Of Voltage Into The Rotor Circuit**

The voltage is injected at slip frequency to the rotor circuit to achieve the desired speed torque characteristic. If the voltage injected opposes the rotor voltage, the effective rotor current decreases, which instantly affects the torque. The reduced torque cannot drive the load. The rotor speed decreases to a value which ensures sufficient induced rotor voltage and hence rotor current to drive the load. If on the other hand, the injected voltage aids the rotor voltage, it results in an increased rotor current. The increased torque developed accelerates the rotor to a speed at which sufficient rotor current flows to drive the load.

- **Variation Of Supply Frequency**

The speed of a synchronously rotating magnetic field is function of supply frequency. Therefore by varying the supply frequency the synchronous speed and hence the rotor speed can be varied. To avoid saturation due to increase in flux at low frequencies, the applied voltage too the motor is also varied, so that the flux remains constant at its rated value at all frequencies. To achieve this a simple methods is to vary both voltage and frequency. So that v/f is constant. The torque speed curves with constant v/f are depicted in the figure. There is a depletion of torque at low frequencies the motor has reduced torque capability and overload capacity. This is because of dominant effect of stator resistance at low frequencies. The resistance drop becomes appreciable as compared to applied voltage. This causes a depletion of flux, whose constancy can not be maintained at low frequencies. The torque developed with v/f constant is:

$$Td = (pm1/2\pi)\,(vm/f1)^2(f2.xm^2/r2')\,/\,[(r1+(f2/r2f'1)\,(xm^2\text{-}x11x22))]^2+ [(x11+(f2r1/f1r2')/x22)]^2 \qquad (8.9)$$

Where: $x11=xm+x1$

$x22=x2'+xm$

To have the made torque and overload frequencies it is necessary to compensate for the stator drop in order E/f constant, v/f is no longer constant since it increases as the frequency decreases. The torque developed in this case is given be

$$Td = pm1/2\pi(E1/f1)^2(f2r2'/(r2'^2+(2\pi f2l2\sigma')^2)) \qquad (8.10)$$

With v/f (constant)control the starting torque increases with a decreases in frequency up to certain value. Below this value of frequency the starting toque decreases. This effect is similar that achieved by changing the rotor leakage reactance. As the frequency decreases the rotor leakage reactance decreases. Effectively an increase in rotor, resistance relative to leakage resistance takes place. Therefore, the starting torque increases up to certain frequency, where the rotor leakage reactance equals the rotor resistance. I the frequency is decreased further the starting torque decreases.

The variation of the starting torque is shown in the figure. However with constant E/f control the starting torque increases as the frequency decreases up to a value decided by the parameters. If the frequency is further decreased the starting torque decreases. The acceleration may be achieved at constant torque and armature current by varying the stator r frequency from a low value, keeping E/f constant.

Speed torque characteristics above the base characteristics can be obtained by increasing supply frequency beyond the rated value. The flux in the motor decreases, since the voltage cannot be increased the rated value. The motor is said to be operating in ht efflux weakening mode.

The torque speed curves runs parallel to each other at all frequencies .They extend to second quadrant, showing that he regeneration is possible. The starting of motor can be easily accomplished using **Variable Voltage Variable Frequency Supply.** The decrease in starting current, giving the reasonably good accelerating torque at a good power factor even with low resistance cage motors.

- **Voltage Boost Required At Low Frequencies**

The effect of rs at low values of frequency f cannot be neglected, even if sf1 is small. In induction motor of normal design, 2πflr is negligible comparison to Rs(f/fsl) in the equivalent circuit. Of figure if Φag is constant eag varies linearly with f, Im is also constant. Therefore, the additional voltage required due to Lls is also proportional to the operating frequency f, the additional voltage required to compensate for the voltage drop in Rs to keep Φag constant, a much higher percentage voltage boost is required at low operating frequencies due to the voltage droop across Rs. At large values of f, the voltage drop across Rs can be neglected in comparison to E_{ag}.

- **(Below rated speed) Constant torque region.**

The stator voltage magnitude is decreased approximately in proportion to frequency from its rated value down to very low values. If flux is maintained constant, the motor can deliver its rated torque (on a continuous base). This region is called the **constant torque region.** Here f_{sl} remains constant at its full load (rated) value with delivering the rated torque. At the constant rated torque the power loss in the rotor resistance is also constant. However in practice getting rid of this rotor heat due to power loss become problem at low speed due to reduced cooling therefore, unless the motor has the constant speed fan or is designed to be totally enclose d and non ventilated, the torque capability drops off at very low speeds. It should be noted that this is of no concern in centrifugal l9oads where the torque requirement is very low a low speeds.

- **(Beyond Rated Speed) Constant Power Region**

By increasing the stator frequency at its nominal value it is possible to increase the motor speed beyond the rated speed. In most adjustable speed drive applications, the motor voltage is not exceeded beyond its rated value. By keeping vs at its rated value, increasing the frequency f results in reduced vs/f and hence, a reduced flux in this region, which results in torque speed curves whose slopes are proportional to $(1/f)^2$ at higher than rated frequencies. In practice, the motor can deliver higher than its rated power by noting that:

  - Im goes down as a result of decreased flux. When Is equals to its rated value allows a higher value of Ir and higher torque and power.
  - Since im is decrease, core losses are reduced.and better cooling at higher speeds.

- **High Speed Operation**

Beyond a speed somewhere in the range of 1.5 to 2 times the rated speed, flux reduces some much that the motor approaches its pullout torque. At still higher speeds the motor can deliver only a fixed percentage of the pull out torques, the torque capability declines as:

$$T_{em,max} \approx k/f^2 \qquad (8.11)$$

Both the torque and motor current decline with speed, keeping vs constant, the motor torque in this region is not limited by the current handling capability of the motor since the current at the limit is less than its rated value and declines with speed. Rather it is limited by the maximum torque produced by the motor.

- **Higher Voltage Operation**

In most motors, the voltages insulation level is much higher than the specified rated voltage of the motor. Therefore, by means of proper solid state power source, it is possible to apply a higher than rated voltage at speeds above the rated speed of motor. This is particularly easy to see in the case of dual voltage motor.

8.3 DYNAMIC d-q MODEL

In an adjustable-speed drive, the machine normally constitutes an element within a feedback loop, and therefore its transient behavior has to be taken into consideration. Besides, high-performance drive control, such as vector or field-oriented control is based on the dynamic d-q model of the machine.

The dynamic performance of an ac machine is somewhat complex because the three phase rotor windings move with respect to the three-phase stator windings. It can be looked on as a transformer with a moving secondary, where the coupling coefficients between the stator and rotor phases change continuously with the change of rotor position θr. The machine model can be described by the differential equations with time varying mutual inductances, but such a model tends to be very complex. Three phase machine can be represented by an equivalent tow-phase machine as shown in fig. 1 where $d^S - q^S$ correspond to stator direct and quadrature axes and d^r-q^r corresponds to the rotor direct and quadrature axes. Although it is somewhat simple, the problem of time varying parameters still remains.

R.H.Park in the 1920 proposed a new theory of electric machine analysis, he replaced the variables(Voltages, currents, and flux linkages) associated with the stator windings of a synchronous machine with variables associated with fictitious windings rotating with the rotor at synchronous speed. In fact, it was shown later by Krause and Thomas that time-varying inductances can be eliminated by referring the stator and rotor variables to a common reference frame which may rotate at any speed.

8.3.1 Axes Transformation

Consider a symmetrical three-phase induction machine with stationary as-bs-cs axes at $2\pi/3$-angle apart, as shown in fig 8.2. Goal is to transform three-phase stationary reference frame(as-bs-cs) variables into two-phase stationary reference frame($d^S - q^S$) variables and the transform these to synchronously rotating reference frame(d^e-q^e), and vice versa.

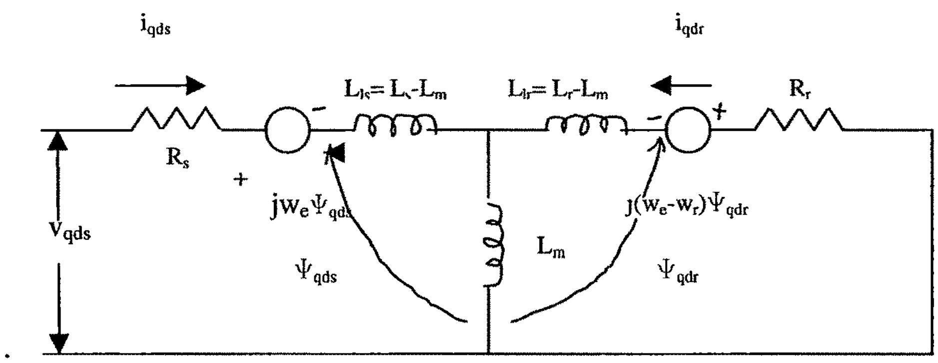


Fig 8.2 Equivalent two phase machine

Assume that the d^S - q^S axes are oriented at θ angle,as shown in figure 3.The voltages $V_{ds}{}^S$ and $V_{qs}{}^S$ can be resolved into as – bs- cs components and can be represented in the matrix form as

$$V_{as} = V_{qs}{}^s \cos\theta + V_{ds}{}^s \sin\theta$$

$$V_{bs} = V_{qs}{}^s \cos(\theta\text{-}120) + V_{ds}{}^s \sin(\theta - 120)$$

$$V_{cs} = V_{qs}{}^s \cos(\theta + 120) + V_{ds}{}^s \sin(\theta + 120) \qquad (8.12)$$

The corresponding inverse rotation

$$V_{qs}{}^s = 2/3(\cos\theta\, V_{qs} + \cos(\theta\text{-}120)\, V_{bs} + \cos(\theta + 120)\, V_{cs}$$

$$V_{ds}{}^s = 2/3(\sin\theta\, V_{qs} + \sin(\theta\text{-}120)\, V_{bs} + \sin(\theta + 120)\, V_{cs} \qquad (8.13)$$

It is convenient to set $\theta = 0$, so that the q^S axis is aligned with the as-axis. the transformation relations can be simplified as

$$V_{as} = V_{qs}{}^s;$$
$$V_{bs} = -½\, V_{qs}{}^s - \sqrt{3}/2\, V_{ds}{}^s$$
$$V_{cs} = -1/2\, V_{qs}{}^s + \sqrt{3}/2\, V_{ds}{}^s \qquad (8.14)$$

and inversely

$$V_{qs}{}^s = 2/3 V_{as} - 1/3\, V_{bs} - 1/3\, V_{cs} = V_{as};$$
$$V_{ds}{}^s = -1/\sqrt{3}\, V_{bs} + 1/\sqrt{3}\, V_{cs} \qquad (8.15)$$

Fig 8.3 shows the synchronously rotating d^e-q^e axes, which rotate at synchronous speed w_e with respect to the d^S - q^S axes and the angle $\theta_e = w_e t$ .The two phase d^S - q^S windings are transformed into the hypothetical windings mounted on the d^e-q^e axes. The voltages on the d^S - q^S axes can be converted (or resolved)into the d^e-q^e frame as follows:

$$V_{qs} = V_{qs}^{\ s}\cos\theta_e - V_{ds}^{\ s}\sin\theta_e$$
$$V_{ds} = V_{qs}^{\ s}\sin\theta_e + V_{ds}^{\ s}\cos\theta_e \qquad (8.16)$$

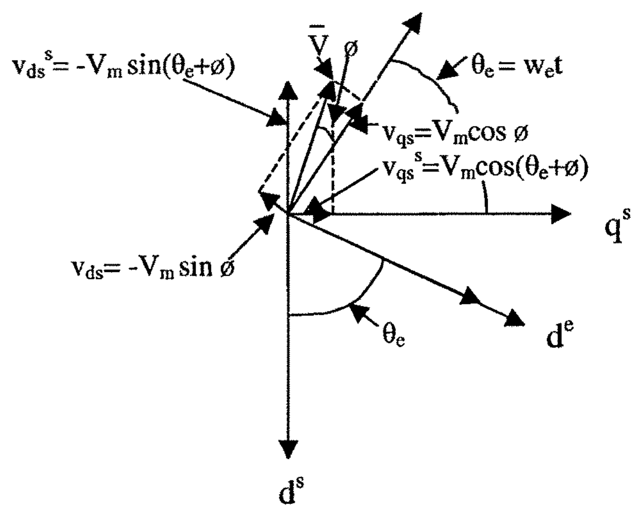


Fig 8.4: Stationary frame d^s-q^s to synchronously rotating frame d^e-q^e transformation.

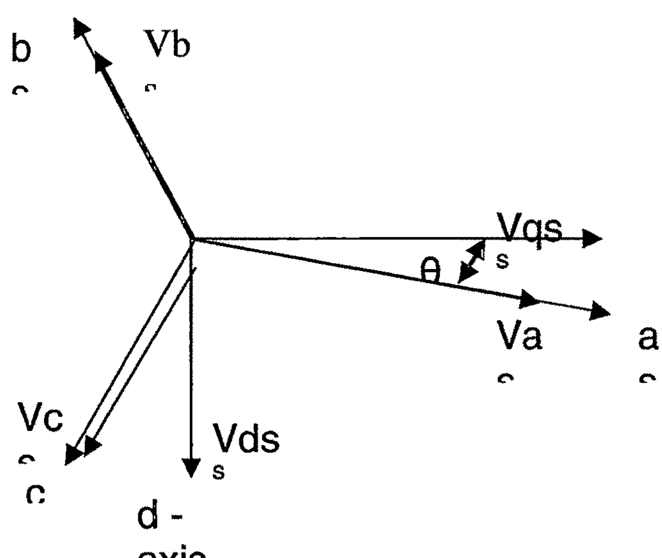


Fig 8.3 Stationary a-b-c to ds-qs axes transformation

For convenience the superscript e has been dropped from now on from the synchronously rotating frame parameters. Resolving the rotating frame parameters into a stationary frame, the relations are

$V_{qs}^{s} = V_{qs} \cos\theta_e + V_{ds} \sin\theta_e$
$V_{ds}^{s} = -V_{qs} \sin\theta_e + V_{ds} \cos\theta_e$ (8.17)

Assume that the three-phase stator voltages are sinusoidal and balanced and are given by:

$V_{as} = V_m\cos(w_e t + \Phi)$
$V_{bs} = V_m\cos(w_e t - 2\prod/3 + \Phi)$
$V_{cs} = V_m\cos(w_e t + 2\prod/3 + \Phi)$ (8.18)

Substituting eqs (8.17) in (8.18) yields

$V_{qs}^{s} = V_m\cos(w_e t + \Phi)$
$V_{ds}^{s} = -V_m\cos(w_e t + \Phi)$ (8.19)

Hence:

$V_{qs} = V_m\cos\Phi$
$V_{ds} = -V_m\sin\Phi$ (8.20)

Equations: V_{qs}^{s} and V_{ds}^{s} are balanced ,two-phase voltages of equal peak values and the latter is at pi/2 angle phase lead with respect to the other component. Sinusoidal variables in a stationary frame appear as dc quantities in a synchronously rotating reference frame. Note that the stator variables are not necessarily balanced sinusoidal waves.

8.4 VECTOR (FIELD-ORIENTED) CONTROL

Control schemes have been developed which provide a feasible approach of speed control to induction motors (*Blaschke*, 1972). The equation of motion describing the steady state behavior of an induction motor are highly nonlinear, time varying and coupled (*Vas*, 1990).

In most applications, speed sensors are necessary and essential in the speed control loop. However, sensors have several disadvantages in terms of drive cost, reliability, and noise immunity.

Various approaches have been proposed for estimating speed using some electric parameters, such as current, voltage, frequency, and flux. They are based on a combination of state estimation theory and vector control theory known as speed sensorless motor control. (Holtz, 1993 ; Hurst, 1994).

Hasse and *Blaschke* developed a vector control theory to simplify the structure of speed control used to drive like DC motors by using coordinate transformations. In recent years, the vector control theory has become more feasible due to progress in development of electronics techniques and high speed microprocessors. *R. H. Park* proposed a new theory of electric machine analysis by replacing the variables (voltages, currents, and flux linkages) associated with the stator windings of a synchronous machine with variables associated with fictitious windings rotating with the rotor at synchronous speed. It was shown later by *Krause* and *Thomas* that time-varying inductances can be eliminated by referring the stator and rotor variables to a common reference frame which may rotate at any speed. In an adjustable-speed drive, the machine normally constitutes an element within a feedback loop.

However, the algorithm of vector control theory requires manipulation of the electric parameters of the motor so that the governing equations in rectangular coordinates cán be developed, prior knowledge of the state equations is necessary when the estimation theory is used to estimate the speed precisely. However, the values of the electric parameters may deviate from the designated values due to changes in the working environment, temperature, speed, external load and noise.

The equations of motion of an induction motor, which are converted by means of

vector control to the type of DC motor control, may be not suitable due to the same reasons, such as changes in the working environment, etc. as mentioned above. Consequently, these unpredicted factors make the actual behavior of a sensorless control motor non-linear and hard to describe. The accuracy will improve if this non-linearity can be governed using other methods in practical applications.

Scalar control techniques of voltage-fed and current-fed inverter drives are somewhat simple to implement, but the inherent coupling effect i.e. both torque and flux are functions of voltage and current and frequency gives sluggish response and the high-order system effect makes it prone to instability. The problem can be solved by vector or field-oriented control. The vector or field-oriented control is based on the dynamic d-q model of the machine.

Vector control is applicable to both induction and synchronous motor drives. The modern sensor less vector control and the corresponding feedback signal processing are complex and needs DSP or micro controllers. Vector control should assure the correct orientation and equality of command and actual currents.

Scalar control techniques of voltage-fed and current-fed inverter drives are somewhat simple to implement ,but the inherent coupling effect i.e. both torque and flux are functions of voltage and current and frequency gives sluggish response and the system is easily prone to instability because of a high-order system effect.

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

8.4.1 DC DRIVE ANALOGY

Ideally, a vector-controlled induction motor drive operates like a separately excited dc motor drive, figure explains this analogy. In a dc machine,neglecting the armature reaction effect and field saturation, the developed torque is given by

$$T_e = k_t' I_a I_f \qquad \textbf{(8.21)}$$

Where I_a = armature current and I_f = field current.

The construction of a dc machine is such that the field flux Ψf produced by the current I_f is perpendicular to the armature flux Ψa , which is produced by the armature current I_a .These space vectors, which are stationary in space,are orthogonal or decoupled in nature. This means that when torque is controlled by controlling the current I_a , the flux Ψf is not affected and we get the fast transient response and high torque/ampere ratio with the rated Ψf. Because of decoupling, when the field current If is controlled, it affects the field flux Ψf only, but not the Ψa flux. Because of the inherent coupling problem, an induction motor cannot generally give such fast response.

DC machine-like performance can also be extended to an induction motor if the machine control is considered in a synchronously rotating reference frame (d^e-q^e), where the sinusoidal variables appear as dc quantities in steady state. In figure, the induction motor with the inverter and vector control in the front end is shown with two control current inputs, i_{ds}* and i_{qs}*. These currents are the direct axis component and quadrature axis component of the stator current, respectively, in a synchronously rotating reference frame. With vector control, i_{ds} is analogous to field current If and i_{qs} is analogous to armature current Ia of a dc machine. Therefore, the torque can be expressed as:

$$T_e = K_t \Psi_r i_{qs}$$
$$T_e = K_t' i_{ds} i_{qs} \qquad (8.22)$$

Where Ψ_r = absolute Ψ_r is the peak value of the sinusoidal space vector. This dc machine like performance is only possible if i_{ds} is oriented in the direction of flux and i_{qs} is established perpendicular to it, as shown by the space-vector diagram on the right of figure.

This means the when i_{qs}* is controlled, it affects the actual i_{qs} current only, but does not affect the flux Ψ_r. Similarly, when i_{ds}*is controlled, it controls the flux only and does not affect the i_{qs} component of current.

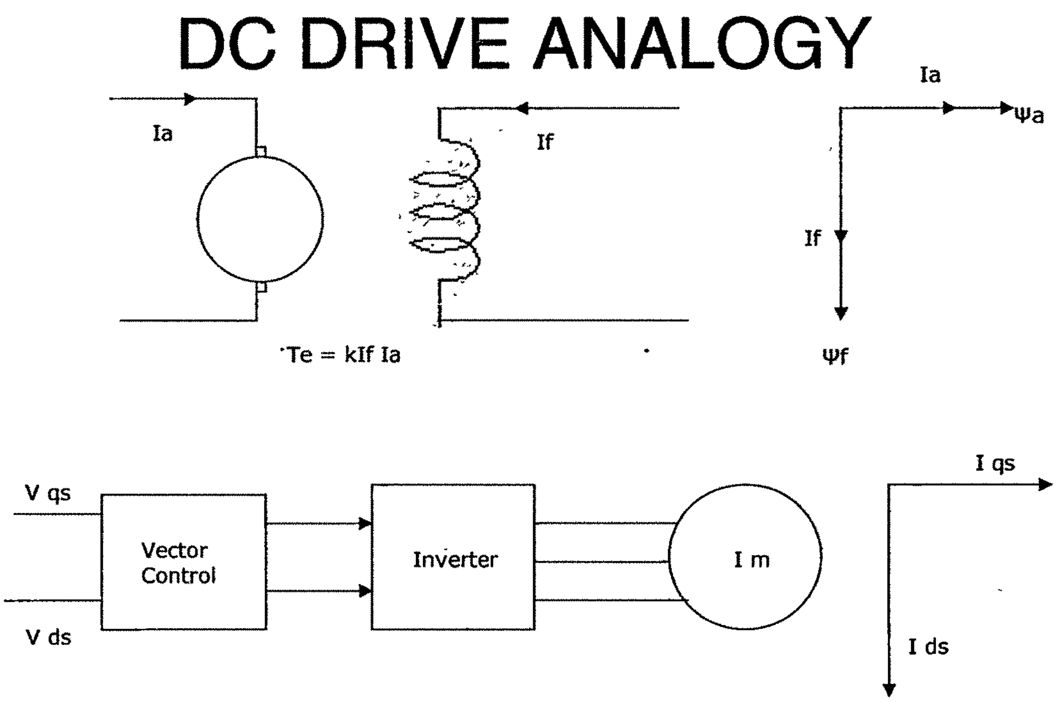


Fig 8.5 Separately excited dc motor, vector-controlled induction motor

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

8.4.2 Principles of Vector Control

The fundamentals of vector control implementation can be explained with the help of fig 8.6, where the machine model is represented in a synchronously rotating reference frame. The inverter is omitted from the figure, assuming that it has unity current gain, that is, it generates currents ia, ib, ic as dictated by the corresponding command currents i_a*, i_b*, i_c* from the controller .A machine model with internal conversions is shown on the right .The mahine terminal phase currents i_a, i_b, i_c are converted to i_{ds}^s and i_{qs}^s components by 3Φ/2Φ transformation. These are then converted to synchronously rotating fame by the unit vector components $\cos\theta_e$ and $\sin\theta_e$ before applying them to the d^e-q^e machine model as shown. The controller makes two stages of inverse transformation, as shown, so that the control currents i_{ds}* and i_{qs}* correspond to the machine currents i_{ds} and i_{qs} respectively. In addition, the unit vector assures correct alignment of i_{ds} current with the flux vector Ψ_r and i_{qs} perpendicular to it as shown. Note that the transformation and inverse transformation including the inverter ideally do not incorporate any dynamics, and therefore, the response to i_{qs} and i_{ds} is instantaneous .

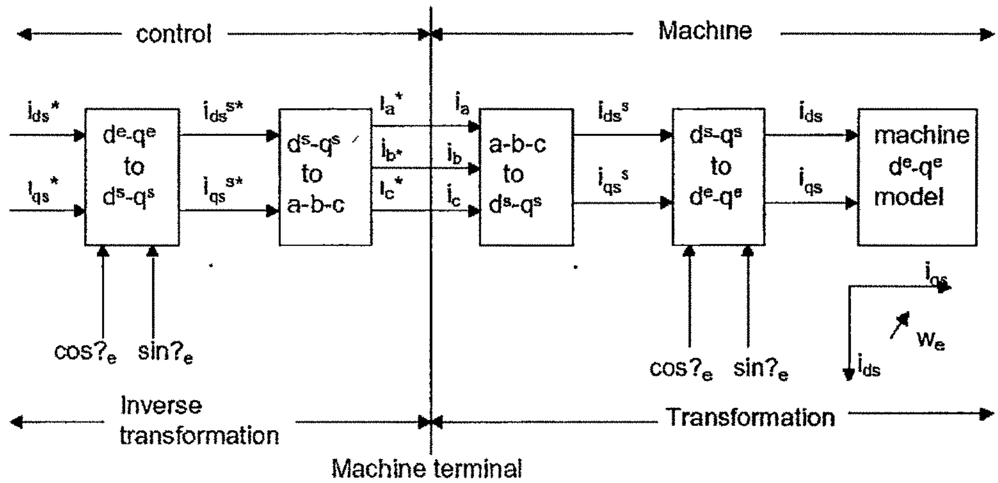


Fig 8.6 Vector control with Machine de-qe model

- **SALIENT features**

- The frequency ω_e of the drive is not directly controlled as in scalar control. The machine is essentially "self-controlled," where the frequency as well as the phase are controlled indirectly with the help of the unit vector.

- There is no fear of an instability problem by crossing the operating point beyond the breakdown torque T_{em} as in a scalar control. Limiting the total $I_s(\sqrt{i_{ds}^2} + i_{qs}^2)$ within the safe limit automatically limits operation within the stable region. The transient response will be fast and dc machine-like because torque control by i_{qs} does not affect the flux. However, note that ideal vector control is impossible in practice because of delays in converter and signal processing and the parameter variation effect.

- Like a dc machine, speed control is possible in four quadrants without any additional control elements (like phase sequence reversing). In forward motoring condition, if the torque Te is negative, the drive initially goes into regenerative braking mode, which slows down the speed. At zero speed, the phase sequence of the unit vector automatically reverse, giving reverse motoring operation.

8.5 Indirect (feed forward) vector control

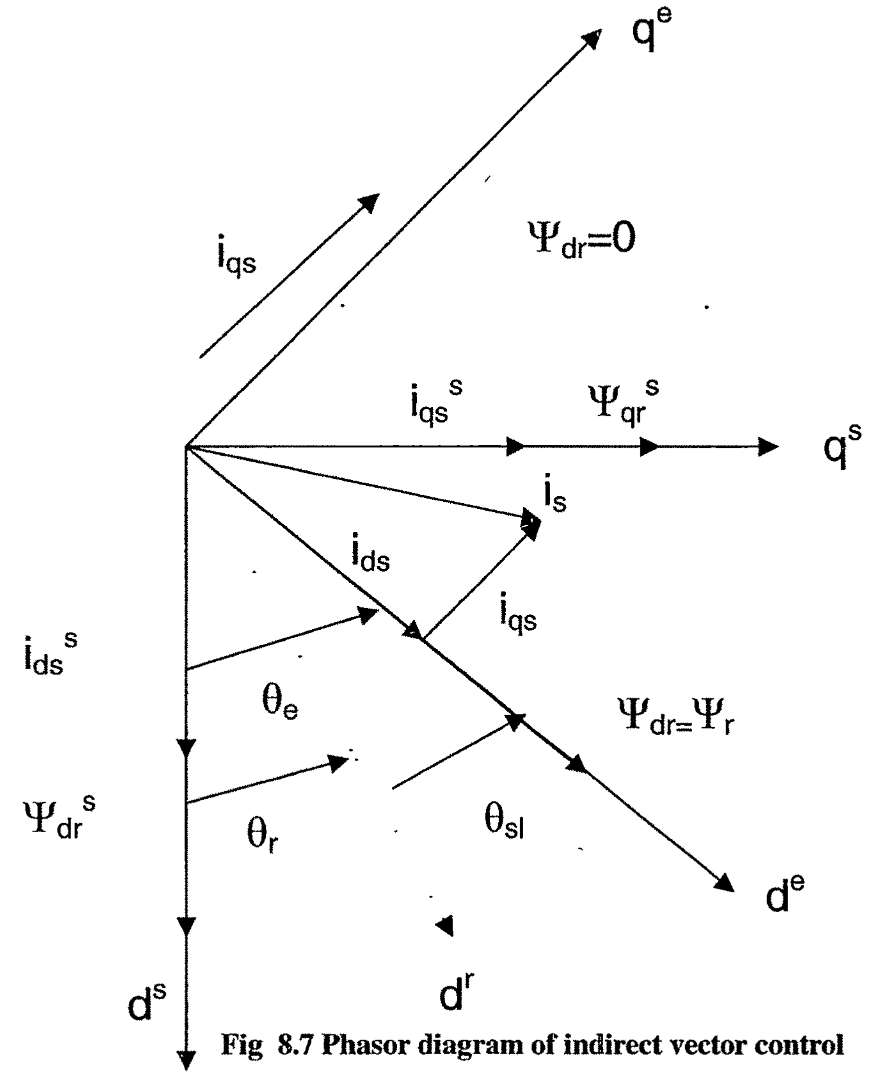


Fig 8.7 Phasor diagram of indirect vector control

The indirect vector control method is essentially the same as direct vector control, except the unit vector signals are generated in feed forward manner. Indirect vector control is very popular in industrial applications. Figure explains the fundamental principle of indirect vector control with the help of a phasor diagram. The d^s-q^s axes are fixed on the stator , but the d^r-q^r axes, which are fixed on the rotor , are moving at spped wr as shown. Synchronously rotating axes d^e-q^e are rotating ahead of the d^r-q^r axes by the positive slip angle θ_{sl} corresponding to slip frequency wsl. Since the rotor pole is directed on the de axis and w_e=w_r+ w_{sl}, so

$$\theta_e = \theta_r + \theta_{sl} \qquad (8.23)$$

Note that the rotor pole position is not absolute, but is slipping with respect to the rotor at frequency w_{sl}. The phasor diagram suggests that for decoupling control, the stator flux component of current ids should be aligned on the de axis, and the torque component of current i_{qs} should be on the q_e axis, as shown.

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

$$d\Psi_{dr}/dt + R_r i_{dr} - (w_e - w_r)\Psi_{qr} = 0 \quad (8.24)$$
$$d\Psi_{qr}/dt + R_r i_{qr} + (w_e - w_r)\Psi_{dr} = 0 \quad (8.25)$$

The rotor flux linkage expressions can be given as

$$\Psi_{dr} = L_r i_{dr} + L_m i_{ds} \quad (8.26)$$
$$\Psi_{qr} = L_r i_{qr} + L_m i_{qs} \quad (8.27)$$

From the above equations, we can write

$$i_{dr} = 1/L_r\, \Psi_{dr} - L_m/L_r\, i_{ds} \quad (8.28)$$
$$i_{qr} = 1/L_r\, \Psi_{qr} - L_m/L_r\, i_{qs} \quad (8.29)$$

The rotor currents can be eliminated with the help of equations (8.28, 8.29) as

$$d\Psi_{dr}/dt + (R_r/L_r)\Psi_{dr} - (L_m/L_r)R_r i_{ds} - w_{sl}\Psi_{qr} = 0 \quad (8.30)$$
$$d\Psi_{qr}/dt + (R_r/L_r)\Psi_{qr} - (L_m/L_r) R_r i_{qs} -+ w_{sl}\Psi_{dr} = 0 \quad (8.31)$$

where: $w_{sl} = w_e - w_r$ has been substituted.

For decoupling control, it is desirable that

$$\Psi_{qr} = 0 \ ; \text{hence: } d\Psi_{qr}/dt = 0 \quad (8.33)$$

so that the total rotor flux Ψ_r is directed on the de axis. Using above conditions:

$$(L_r/R_r)d\Psi_r^*/dt + \Psi_r^* = L_m i_{ds} \quad (8.34)$$
$$w_{sl} = (L_m R_r/\Psi_r L_r)\, i_{qs} \quad (8.35)$$

Where: $\Psi_r^* = \Psi_{dr}$ has been substituted.

If rotor flux Ψ_r^* = constant, $\Psi_r^* = L_m i_{ds}$ (8.36)

In other words, the rotor flux is directly proportional to current ids in steady state.
To implement the indirect vector control strategy, synchronous current control voltage PWM is used to generates the torque component of current i_{ds}^* , as usual .The flux component of current i_{ds}^* for the desired rotor flux Ψ_r^* is determined and maintained constant here in the open loop manner for simplicity. The variation of magnetizing inductance Lm will cause some drift in the flux . The slip frequency w_{sl}^* is generated from i_{qs}^* in feed-forward manner to satisfy the phasor diagram in fig 8.7. The corresponding expression of slip gain K_s is given as

$$K_s = w_{sl}^*/i_{qs}^* = (L_m R_r)/(L_r \Psi_r^*) \quad (8.37)$$

Signal w_{sl}^* is added with speed signal w_r to generate frequency signal w_e. The unit vector signals $\cos\theta_e$ and $\sin\theta_e$ are then generated from w_e by integration and look-up table.

The 2Φ/3Φ transformation are the same .The speed signal from an incremental-position encoder is mandatory in indirect vector control because the slip signal locates the pole with respect to the rotor d^r axis in feed forward manner, which is moving a t speed w_r. An absolute pole position on the rotor is not required in this case like a synchronous motor. If the polarity of i_{qs}* becomes negative for negative torque, the phasor i_{qs} in fig 8.8 will be reversed and w_{sl} will be negative, which will shift the rotor pole position de axis below the d^r axis.

The speed control range in indirect vector control can easily be extended from stand-still to the field-weakening region. In the constant torque region, the flux is constant. However, in the field-weakening region, the flux is programmed such that the inverter always operates in PWM mode.

Command currents i_{ds}* and i_{qs}* in the vector control are compared with the respective ids and i_{qs} currents generated by the transformation of phase currents with the help of unit vector. The respective errors generate the voltage command signals v_{ds}* and v_{qs}* through the P-I compensators, as shown. These voltage commands are then converted into stationary frame phase voltages. The synchronous frame current control with a P-I controller assure amplitude and phase tracking of currents, even when the PWM controller goes in to over-modulation range.

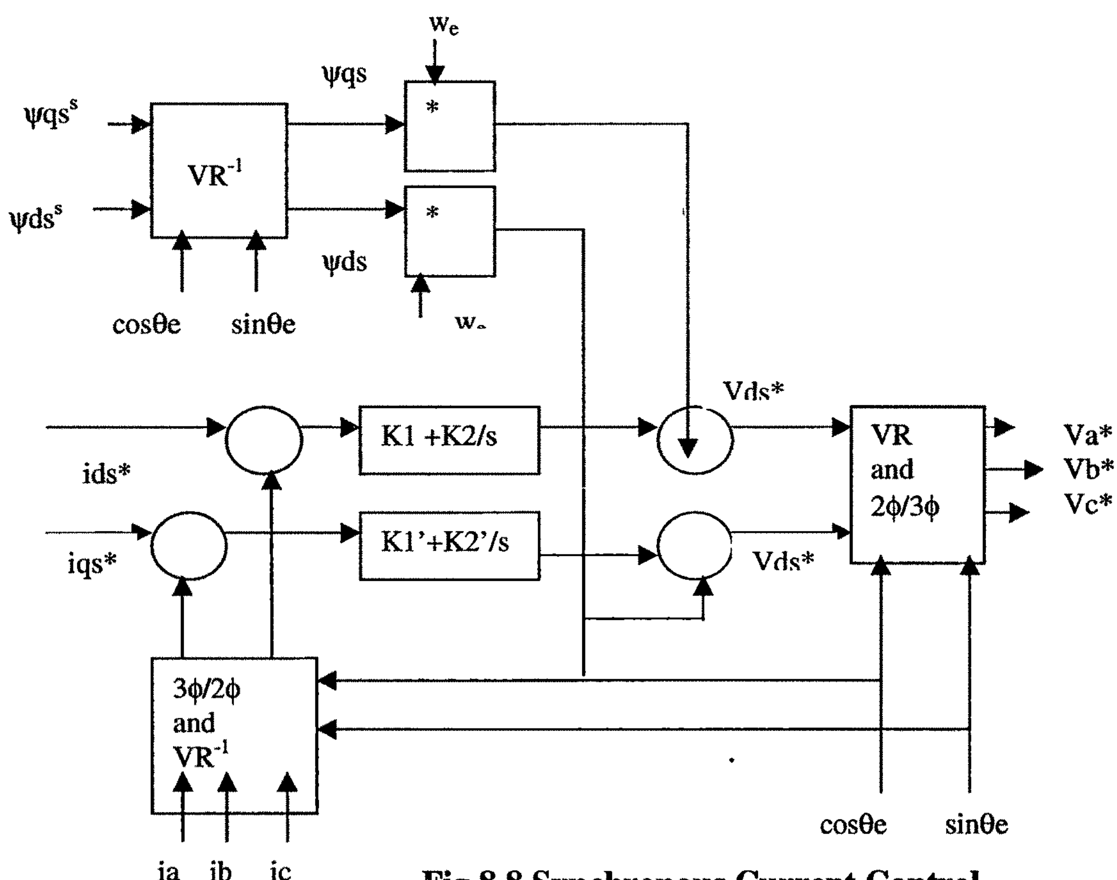


Fig 8.8 Synchronous Current Control

8.5.1 Synchronously rotating reference frame-dynamic model (Kron equation)

For the 2-Φ machine, we need to represent both d^s–q^s and d^r–q^r circuits and their variables in a synchronously rotating d^e–q^e frame. Fig 8.9 shows the d^e–q^e dynamic model equivalent circuits. A special advantage of the d^e–q^e dynamic model of the machine is that all the sinusoidal variables in stationary frame appear as dc quantities in synchronous frame.

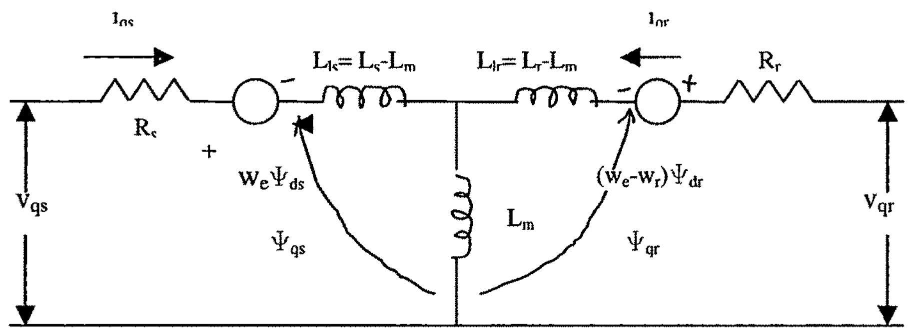


Fig 8.9(a): Dynamic d^e-q^e equivalent circuit of machine q^e-axis circuit.

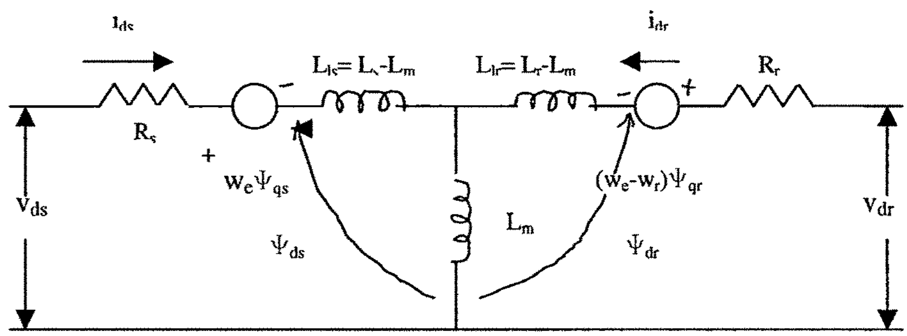


Fig 8.9(b): Dynamic d^e-q^e equivalent circuit of machine d^e-axis circuit.

The flux linkage expressions in terms of the currents can be written from fig (8.9) as follows:

$Y_{qs} = L_{ls}i_{qs} + L_m(i_{qs} + i_{qr})$

$Y_{qr} = L_{lr}i_{qr} + L_m(i_{qs} + i_{qr})$

$Y_{qm} = L_m(i_{qs} + i_{qr})$

$Y_{ds} = L_{ls}i_{ds} + L_m(i_{ds} + i_{dr})$ ----------(8.39)

$Y_{dr} = L_{lr}i_{dr} + L_m(i_{ds} + i_{dr})$ ----------(8.40)

$Y_{dm} = L_m(i_{ds} + i_{dr})$ ----------(8.41)

Combining above expressions the electrical transient model in terms of voltages and currents can be given in matrix form as

$$\begin{bmatrix} v_{qs} \\ v_{ds} \\ v_{qr} \\ v_{qr} \end{bmatrix} = \begin{bmatrix} R_s + SL_s & w_e & SL_m & w_eL_m \\ -w_eL_s & R_s + SL_s & -w_eL_m & SL_m \\ SL_m & (w_e-w_r)L_m & R_r + SL_r & (w_e-w_r)L_r \\ -(w_e-w_r)L_m & SL_m & -(w_e - w_r)L_r & R_r + SL_r \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{qr} \\ i_{dr} \end{bmatrix} \quad \text{-----(8.42)}$$

where S is the Laplace operator. For a single-fed machine, such as a cage motor, $v_{qr} = v_{dr} = 0$. The speed w_r can be related to the torques as

$$T_e = T_L + J\frac{dw_m}{dt} = T_L + \frac{2J}{P}\frac{dw_r}{dt} \quad \text{----------(8.43)}$$

where T_L = load torque,
J = rotor inertia,
w_m = mechanical speed.

The machine model can be represented in complex form. Fig (8.10) shows the complex equivalent circuit in rotating frame where v_{qdr}=0.

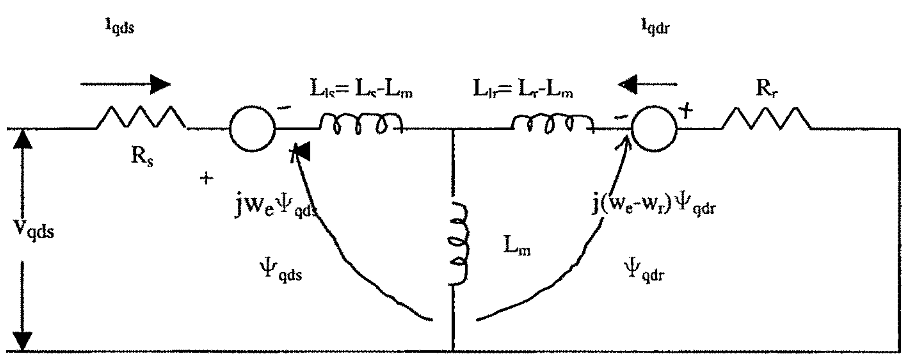


Fig 8.10: Complex synchronous frame dqs equivalent circuit.

Resolving the variables into d^e-q^e components, as shown in fig (8 11)

$$T_e = \frac{3}{2}\left[\frac{P}{2}\right] (\Psi_{dm} i_{qr} - \Psi_{qm} i_{dr}) \quad \text{----------(8.43)}$$

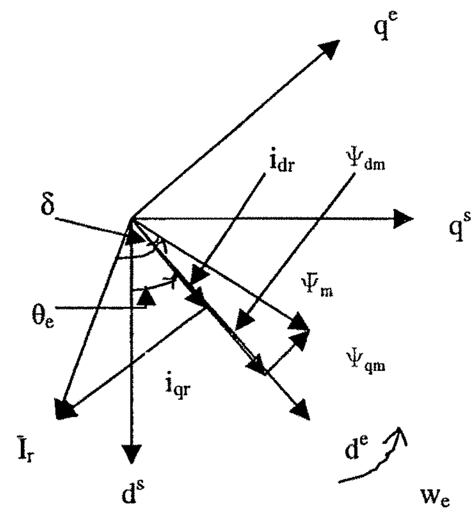


Fig 8.11: Flux and current vectors in d^e-q^e frame.

Several other torque expressions can be derived easily as follows:

$$T_e = \frac{3}{2}\left[\frac{P}{2}\right](\Psi_{ds}i_{qs} - \Psi_{qs}i_{ds}) \qquad \text{----------(8.44)}$$

Equations: (8.42), (8.43) and (8.44) give the complete model of the electro-mechanical dynamics of an induction machine in synchronous frame. Fig (8.12) shows block diagram of the machine model along with input voltage and output current transformations.

8.5.2 Dynamic model state-space equations

The dynamic machine model in state-space form is important for transient analysis, particularly for computer simulation study. The electrical variables in the model can be chosen as fluxes, currents, or a mixture of both. State-space equations of the machine in rotating frame with flux linkages as the main variables are derived. To obtain dynamic machine model in stationary frame, we have to substitute $w_e = 0$.

Let

$$F_{qs} = w_b\Psi_{qs} \qquad \text{----------(8.45)}$$

$$F_{qr} = w_b\Psi_{qr} \qquad \text{----------(8.46)}$$

$$F_{ds} = w_b\Psi_{ds} \qquad \text{----------(8.47)}$$

$$F_{dr} = w_b\Psi_{dr} \qquad \text{----------(8.48)}$$

where w_b = base frequency.

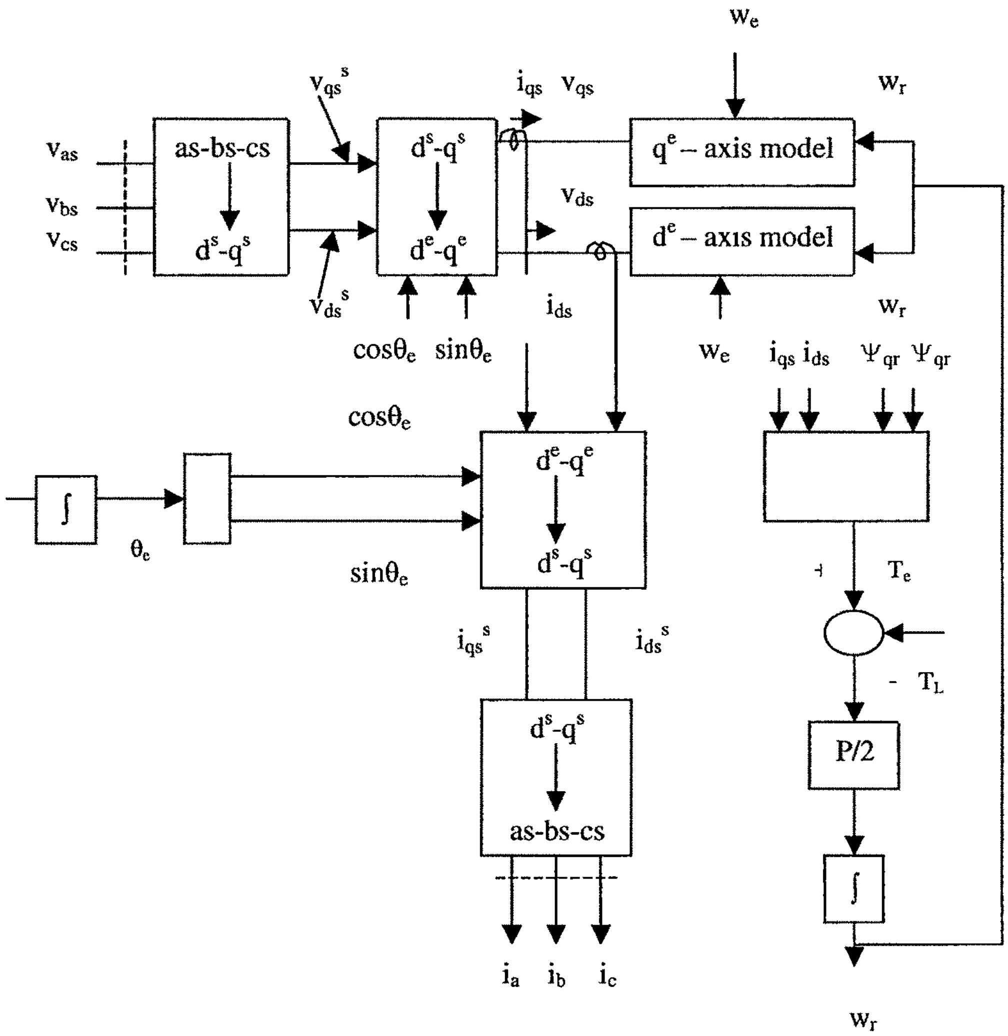


Fig 8.12: Synchronously rotating frame machine model with input voltage and output current transformations.

Using above relations, we can write

$$v_{qs} = R_s i_{qs} + \frac{1}{w_b}\frac{dF_{qs}}{dt} + \frac{w_e}{w_b} F_{ds} \qquad \text{----------(8.49)}$$

$$v_{ds} = R_s i_{ds} + \frac{1}{w_b}\frac{dF_{ds}}{dt} - \frac{w_e}{w_b} F_{qs} \qquad \text{----------(8.50)}$$

$$0 = R_r i_{qr} + \frac{1}{w_b}\frac{dF_{qr}}{dt} + \frac{(w_e - w_r)}{w_b} F_{dr} \qquad \text{----------(8.51)}$$

$$0 = R_r i_{dr} + 1\ dF_{dr} - (w_e - w_r) F_{qr} \qquad \text{----------(8.52)}$$

$$w_b\,dt \qquad\qquad w_b$$

where it is assumed that $v_{qr} = v_{dr} = 0$.

The flux linkage expressions can be written as

$$F_{qs} = w_b\Psi_{qs} = X_{ls}i_{qs} + X_m(i_{qs} + i_{qr}) \qquad \text{----------(8.53)}$$

$$F_{qr} = w_b\Psi_{qr} = X_{lr}i_{qr} + X_m(i_{qs} + i_{qr}) \qquad \text{----------(8.54)}$$

$$F_{qm} = w_b\Psi_{qm} = X_m(i_{qs} + i_{qr}) \qquad \text{----------(8.55)}$$

$$F_{ds} = w_b\Psi_{ds} = X_{ls}i_{ds} + X_m(i_{ds} + i_{dr}) \qquad \text{----------(8.56)}$$

$$F_{dr} = w_b\Psi_{dr} = X_{lr}i_{dr} + X_m(i_{ds} + i_{dr}) \qquad \text{----------(8.57)}$$

$$F_{dm} = w_b\Psi_{dm} = X_m(i_{ds} + i_{dr}) \qquad \text{----------(8.58)}$$

where $X_{ls} = w_bL_{ls}$

$X_{lr} = w_bL_{lr}$

$X_m = w_bL_m$

or

$$F_{qs} = X_{ls}i_{qs} + F_{qm} \qquad \text{----------(8.59)}$$

$$F_{qr} = X_{lr}i_{qr} + F_{qm} \qquad \text{----------(8.60)}$$

$$F_{ds} = X_{ls}i_{ds} + F_{dm} \qquad \text{----------(8.61)}$$

$$F_{dr} = X_{lr}i_{dr} + F_{dm} \qquad \text{----------(8.62)}$$

The currents can be expressed in terms of the flux linkages as:

$$i_{qs} = \frac{F_{qs} - F_{qm}}{X_{ls}} \qquad \text{----------(8.63)}$$

$$i_{qr} = \frac{F_{qr} - F_{qm}}{X_{lr}} \qquad \text{----------(8.64)}$$

$$i_{ds} = \frac{F_{ds} - F_{dm}}{X_{ls}} \qquad \text{----------(8.65)}$$

$$i_{dr} = \frac{F_{dr} - F_{dm}}{X_{lr}} \qquad \text{----------(8.66)}$$

The F_{qm} is given by,

$$F_{qm} = X_m\left[\frac{(F_{qs} - F_{qm})}{X_{ls}} + \frac{(F_{qr} - F_{qm})}{X_{lr}}\right] \qquad \text{----------(8.67)}$$

or

$$F_{qm} = \frac{X^*_{m1}}{X_{ls}} F_{qs} + \frac{X^*_{m1}}{X_{lr}} F_{qr} \qquad \text{----------(8.68)}$$

$$\text{where } X^*_{m1} = \frac{1}{\left[\frac{1}{X_m} + \frac{1}{X_{ls}} + \frac{1}{X_{lr}}\right]} \qquad \text{----------(8.69)}$$

Similarly F_{dm} is given by,

$$F_{dm} = \frac{X^*_{m1}}{X_{ls}} F_{ds} + \frac{X^*_{m1}}{X_{lr}} F_{dr} \qquad \text{----------(8.70)}$$

Using current equations into voltage equations

$$v_{ds} = \frac{R_s}{X_{ls}}(F_{qs} - F_{qm}) + \frac{1}{w_b}\frac{dF_{qs}}{dt} - \frac{w_e}{w_b} F_{ds} \qquad \text{----------(8.71)}$$

$$v_{ds} = \frac{R_s}{X_{ls}}(F_{ds} - F_{dm}) + \frac{1}{w_b}\frac{dF_{ds}}{dt} - \frac{w_e}{w} F_{qs} \qquad \text{----------(8.72)}$$

$$0 = \frac{R_r}{X_{lr}}(F_{qr} - F_{qm}) + \frac{1}{w_b}\frac{dF_{qr}}{dt} + \frac{(w_e - w_r)}{w_b} F_{dr} \qquad \text{----------(8.73)}$$

$$0 = \frac{R_r}{X_{lr}}(F_{dr} - F_{dm}) + \frac{1}{w_b}\frac{dF_{qr}}{dt} - \frac{(w_e - w_r)}{w_b} F_{qr} \qquad \text{----------(8.74)}$$

The equations expressed in state-space form are:

$$\frac{dF_{qs}}{dt} = w_b\left[v_{qs} - \frac{w_e}{w_b} F_{ds} - \frac{R_s}{X_{ls}}(F_{qs} - F_{qm})\right] \qquad \text{----------(8.75)}$$

$$\frac{dF_{ds}}{dt} = w_b\left[v_{ds} - \frac{w_e}{w_b} F_{qs} - \frac{R_s}{X_{ls}}(F_{ds} - F_{dm})\right] \qquad \text{----------(8.76)}$$

$$\frac{dF_{qr}}{dt} = -w_b\left[\frac{(w_e - w_r)F_{dr}}{w_b} + \frac{R_r}{X_{lr}}(F_{qr} - F_{qm})\right] \qquad \text{----------(8.77)}$$

$$\frac{dF_{dr}}{dt} = -w_b\left[-\frac{(w_e - w_r)F_{qr}}{w_b} + \frac{R_r}{X_{lr}}(F_{dr} - F_{dm})\right] \qquad \text{----------(8.78)}$$

$$T_e = \frac{3}{2}\left[\frac{P}{2}\right]\frac{1}{w_b}(F_{ds}i_{qs} - F_{qs}i_{ds}) \qquad \text{------(8.79-a)}$$

$$dw_r = \left[\;P\;\right](T_e - T_L) \qquad \text{-------(8.79-b)}$$

$\overline{\text{dt}}$ $\overline{\text{2J}}$

Equations (8.75)-(8.79) describe the complete model in state-space form where F_{qs}, F_{ds}, F_{qr} and F_{dr} are the state variables.

$$\frac{dF_{qs}}{dt} = w_b\left[v_{qs} - \frac{w_e}{w_b}F_{ds} + \frac{R_s}{X_{ls}}\left[\frac{X^*_{ml}}{X_{lr}}F_{qr} + \left[\frac{X^*_{ml}}{X_{ls}} - 1\right]F_{qs}\right]\right] \quad \text{--(8.80)}$$

$$\frac{dF_{ds}}{dt} = w_b\left[v_{ds} + \frac{w_e}{w_b}F_{qs} + \frac{R_s}{X_{ls}}\left[\frac{X^*_{ml}}{X_{lr}}F_{dr} + \left[\frac{X^*_{ml}}{X_{ls}} - 1\right]F_{ds}\right]\right] \quad \text{--(8.81)}$$

$$\frac{dF_{qr}}{dt} = w_b\left[\frac{-(w_e-w_r)F_{dr}}{w_b} - \frac{R_r}{X_{lr}}\left[\frac{X^*_{ml}}{X_{ls}}F_{qs} - \left[\frac{X^*_{ml}}{X_{lr}} - 1\right]F_{qr}\right]\right] \quad \text{--(8.82)}$$

$$\frac{dF_{dr}}{dt} = w_b\left[\frac{(w_e-w_r)F_{qr}}{w_b} - \frac{R_r}{X_{lr}}\left[\frac{X^*_{ml}}{X_{ls}}F_{ds} - \left[\frac{X^*_{ml}}{X_{lr}} - 1\right]F_{dr}\right]\right] \quad \text{--(8.83)}$$

Substituting for F_{qm}, F_{dm} and simplifying state-space model in matrix form is::

$$dx/dt = Ax + Bv \quad \text{where: } x = [F_{qs}\ F_{ds}\ F_{qr}\ F_{dr}]^T \text{ and } v = [vqs\ vds\ 0\ 0\ ]T$$

The system matrices are:

$$A = \begin{bmatrix} \frac{R_s w_b}{X_{ls}}\left[\frac{X^*_{ml}}{X_{ls}} - 1\right] & -w_e & \frac{R_s w_b X^*_{ml}}{X_{ls}X_{lr}} & 0 \\ w_e & \frac{R_s w_b}{X_{ls}}\left[\frac{X^*_{ml}}{X_{ls}} - 1\right] & 0 & \frac{R_s w_b X^*_{ml}}{X_{ls}X_{lr}} \\ \frac{R_r w_b X^*_{ml}}{X_{lr}X_{ls}} & 0 & \frac{R_r w_b}{X_{lr}}\left[\frac{X^*_{ml}}{X_{lr}} - 1\right] & -(w_e-w_r) \\ 0 & \frac{R_r w_b X^*_{ml}}{X_{lr}X_{ls}} & (w_e-w_r) & \frac{R_r w_b}{X_{lr}}\left[\frac{X^*_{ml}}{X_{lr}} - 1\right] \end{bmatrix}$$

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

STATE SPACE MODEL

8.6 SIMULINK Models

Usually, when an electrical machine is simulated in circuit simulators like Pspice, its steady state model is used, but for electrical drive studies, the transient behavior is also important. One advantage of SIMULINK over circuit simulators is the ease ın modeling the transients of electrical machines and drives and to include drive controls in the simulation. As long as the equations are known, any drive or control algorithm can be modeled in Simulink.

Some of the equations recommend using S-functions , which are software source codes for Simulink blocks. This technique does not fully utilıze the power and ease of Simulink because S-function programming knowledge is required to access the model variables. S–functions run faster than discrete Simulink blocks, but Simulink models can be made to run faster using "accelerator" functions or producing stand-alone Simulink models. Both of these require additional expense and can be avoided if the simulation speed is not that critical. Another approach is using the Simulink Power System Block set that can be purchase with Simulink. This block set also makes use of S-functions and is not as easy to work with as the rest of the Simulink blocks.

Simulink Induction motor model is described. With the modular system, each block solves one of the model equations; therefore, unlike black box models, all of the machine parameter is accessible for control and verification purposes.

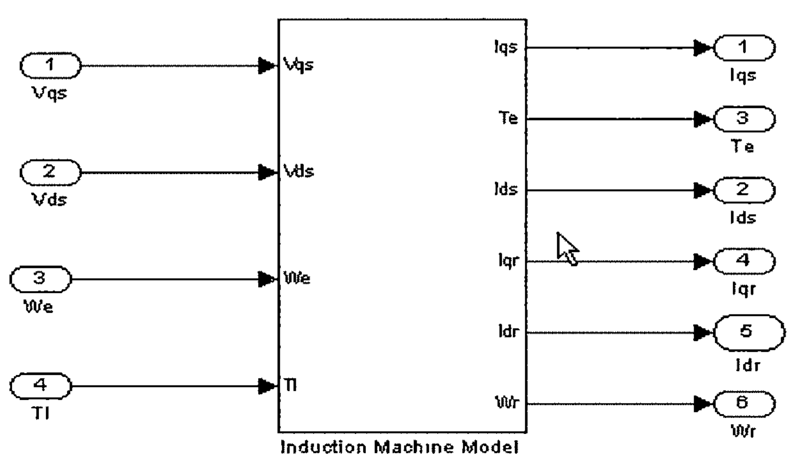


Fig 8.13: SIMULINK implementation of Induction Machine Model.

The simulink set-up for d-q flux linkages model of induction machine is as shown in the fig (8.14). Table 8.1 represents details of each simulink block and relation with the equations. Fig 8.15 depicts the complete simulink setup. Table 8.2 gives details of files used for storing nformations.

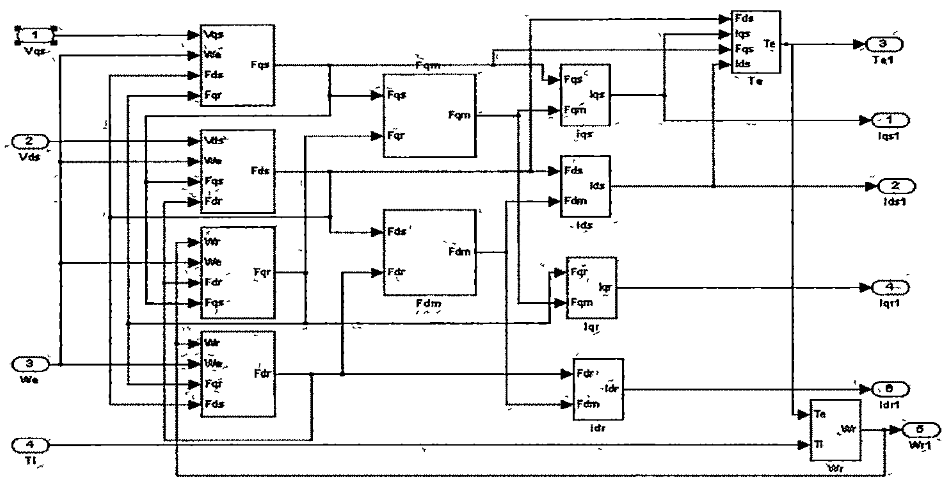


Fig 8.14: SIMULINK set-up of d^e-q^e flux linkage model of Induction motor

BLOCK DIAGRAM	QUANTITY	EQUATION NUMBER
V_{qs}, W_e, F_{ds}, F_{qr} → F_{qs}	F_{qs}	8.75
V_{ds}, W_e, F_{qs}, F_{dr} → F_{ds}	F_{ds}	8.76
W_r, W_e, F_{dr}, F_{qs} → F_{qr}	F_{qr}	8.77
W_r, W_e, F_{qr}, F_{ds} → F_{dr}	F_{dr}	8.78

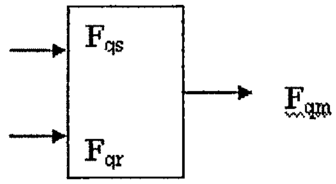
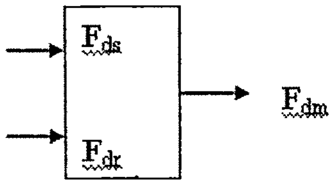
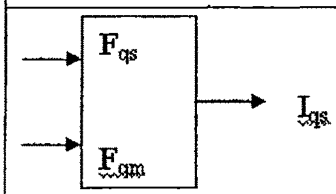
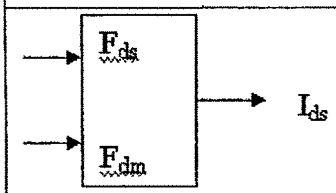
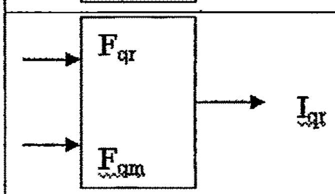
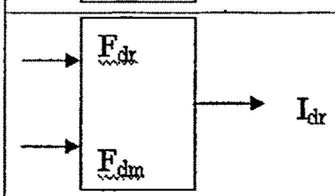
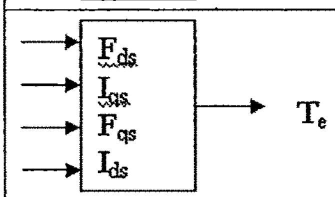
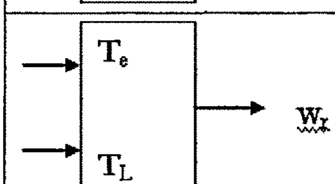
	F_{qm}	8.68
	F_{dm}	8.70
	I_{qs}	8.63
	I_{ds}	8.64
	I_{qr}	8.65
	I_{qs}	8.66
	T_e	8.79-a
	w_r	8.79-b

Table 8.1: Block description of d^e-q^e flux linkage model of induction motor

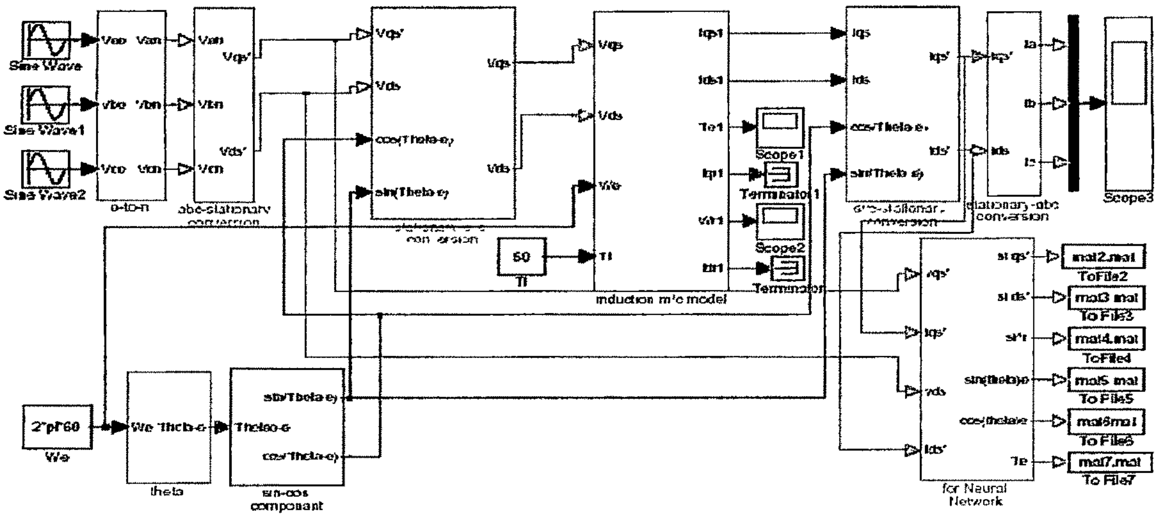


Fig 8.15: Complete Induction machine SIMULINK model.

FILE NAME	PARAMETERS	VARIABLES
Scope1	Motor torque	T_{e1}
Scope2	Rotor speed	w_{r1}
Scope3	Phase currents	I_a,I_b,I_c
mat2.m	q-axis current	I_{qs}
mat3 m	d-axis current	I_{ds}
mat4 m	Rotor current	I_r
mat5.m	Unit vector	$\sin\theta_e$
mat6 m	Unit vector	$\cos\theta_e$
mat7.m	Motor torque	T_e

Table 8.2: Stored Variables (quantities) in different files.

8.7 SIMULATION RESULTS:

- **Initialization:** To simulate the machine in simulink the simulink model has to be initialized first so that it will know all the machine parameters. For this reason, an initialization file assigns values to the machine parameters is formed. This file assigns values to the machine parameter variables in the Simulink model.

  Fig 8 16 shows the initialization file for w 30KW induction machine Before the simulation, this file has to be executed at the Matlab prompt, otherwise Simulink will display an error message. Note that this initialization file is the only machine specific part of the Simulink model that needs to be changed when the machine to be simulated is changed.

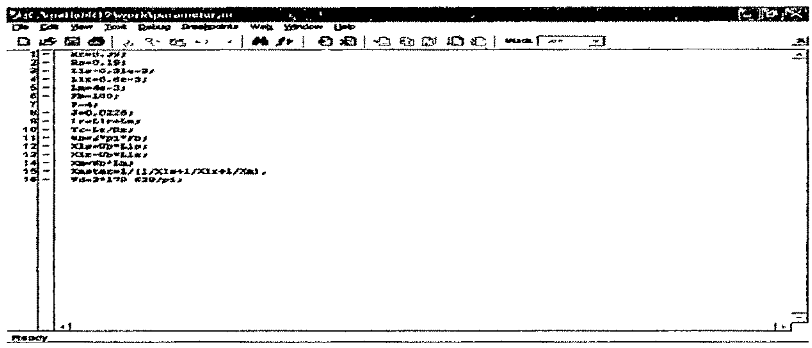
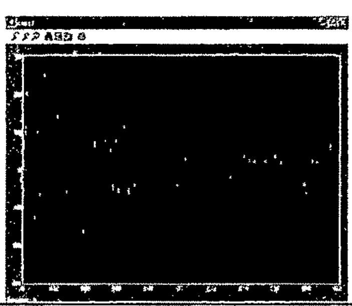
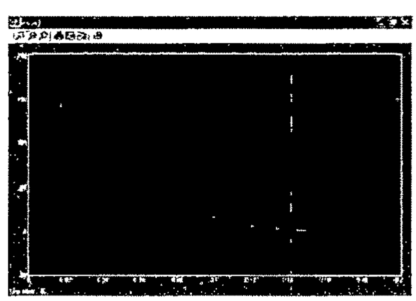
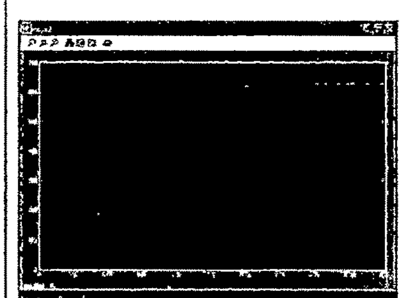


Fig 8.16 Induction machine model initialization file

- **Direct ac start-up:** To test thıs model, the ınduction machıne with the parameters gıven in figure 6 is sımulated by applying 230V three-phase ac voltages at 60HZ with just an inertıa load. Fig 8.17show the three-phase currents, torque and speed of the machine, respectively. The machine accelerates and comes to steady state at 0 14 s with a small slip because of the inertıa load.

		
3φ Motor currents	**Developed torque**	**Rotor speed**
Fig 8.17: Simulation Test Results		

8.7.1 Open-loop constant V/HZ operation

Fig 8.18 shows the implementation of open-loop constant V/Hz control of an inductıon machıne This figure has two new blocks command voltage generator and 3-phse PWM inverter blocks The first one generates the three-phase voltage commands, and it is nothing more than a "syn-abc" block explained earlier. The latter first compares the reference voltage, vref to the command voltages to generate PWM sıgnals for each phase, then uses there signals to drive three Simulink "Swıtch" blocks switching between +Vd/2 and –d/2 (Vd:dc link voltage).

The open-loop constant V/HZ operation is sımulated for 1 2s rampıng up and down the speed command and applying step load torques The results are plotted in Fıg 19 depicts the response of the drive to variation ın speed and the load disturbances.

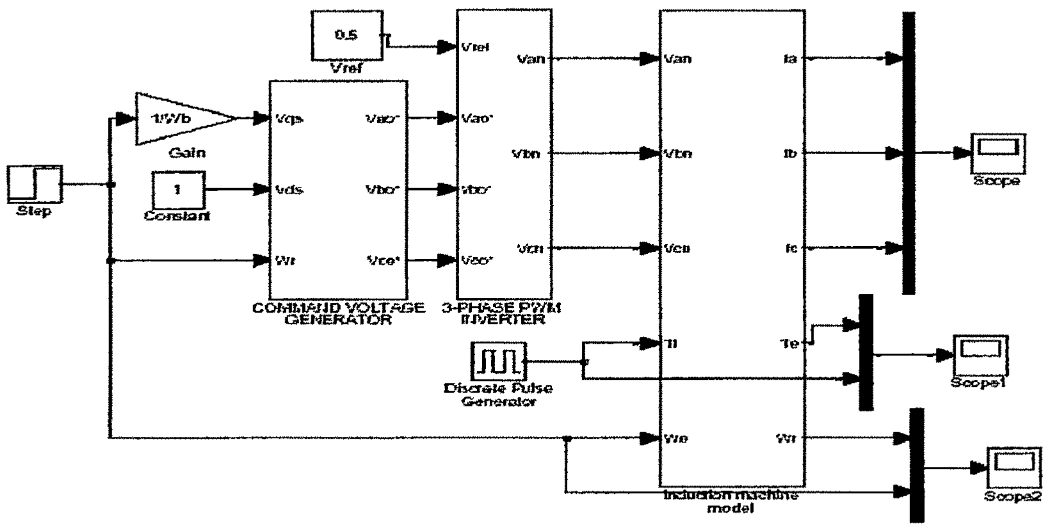


Fig 8.18 V/Hz control- SIMULINK Model

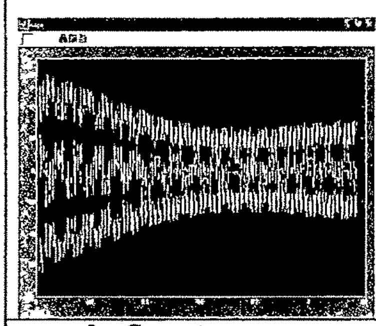
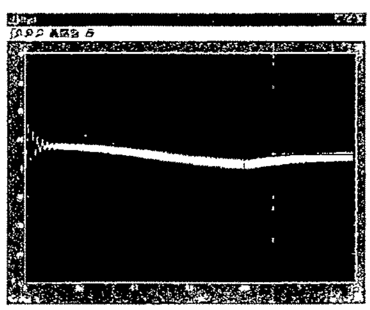
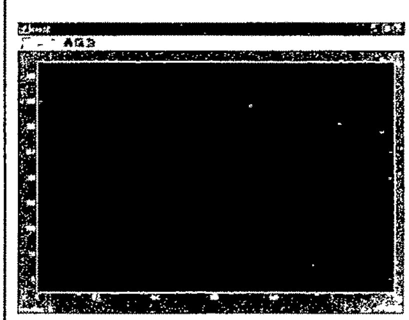
		
3-φ Currents	**Developed torque (Yellow) & load torque (marun)**	**Rotor speed(yellow) & Applied speed (marun)**

Fig 8.19: V/Hz Control

8.7.2 Indirect Vector Control Operation

By addıng a vector control block in fig 8.18 and d-q current feedback, the ınduction machine can be simulated under ındırect vector control The resultıng block diagram ıs shown in fig 8.20, on the other hand, shows the vector control block. Note that, the current feedback for the current controllers comes directly from the induction machine d-q model through Simulink "Goto" and "From" blocks to reduce the clutter of sıgnal lınes

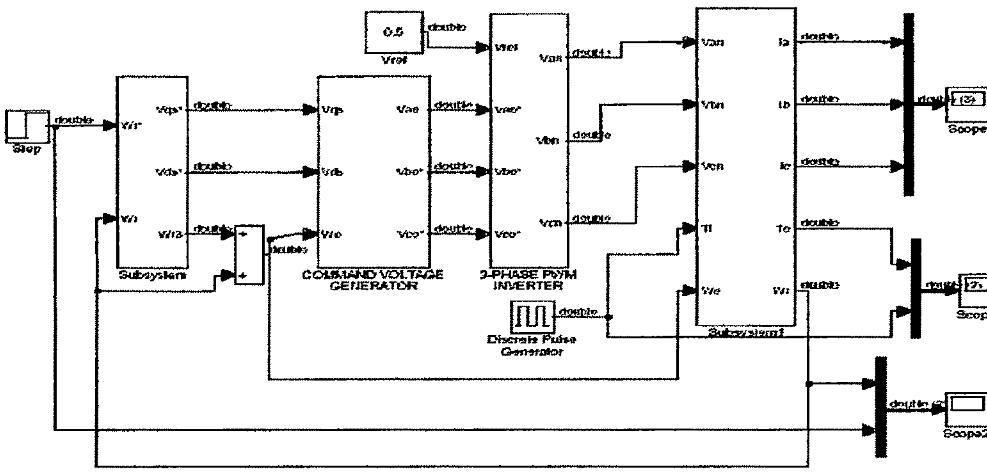
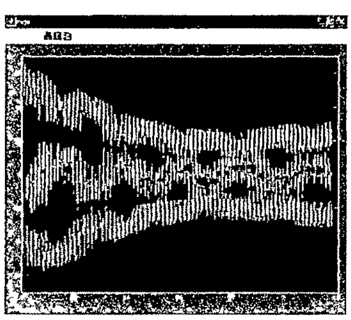
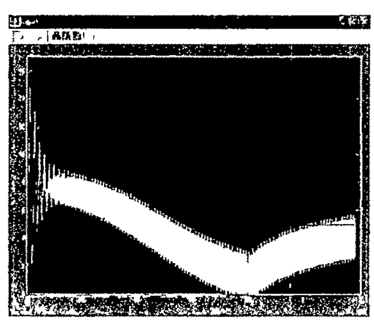
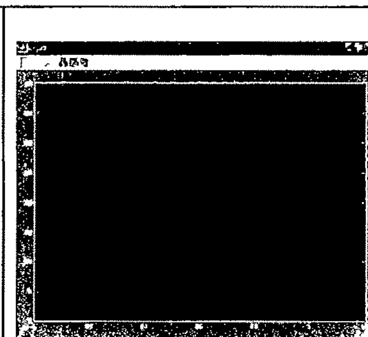


Fig 8.20 Indirect vector control SIMULINK model

The results of the vector control simulation are shown in Fig 8 21 which indicates that the speed tracking and the response to the torque disturbance are excellent

		
IM currents	**Torques:** **load torque** **Developed Torque**	**Synchronous speed &** **Rotor speed**
Fig 8.21 Indirect vector control SIMULINK model: Response		

8.8 ANN Controller: GENERAL DESIGN METHODOLOGY

The design steps for ANN based controller are as follows

1 Analyze the problem and find whether it has sufficient elements for a neural network solution. Consider alternative approaches. A simple DSP/ASIC based direct solution may be satisfactory

2. If the ANN is to represent a static function, then a three-layer feedforward network should be sufficient. For a dynamic function select either a recurrent network or a

time delayed network. Information about the structure and order of the dynamic system is required.

3. Select input nodes equal to the number of input signals and output nodes equal to the number of output signals and a bias source. For a feedforward network, select the initially hidden layer neurons typically mean of input and output nodes.

4. Create an input/output training data table. Capture the data from an experimental plant or simulation results, if possible.

5. Select input scale factor to normalize the input signals and the corresponding output scale factor for denormalization.

6. Select generally a sigmoidal transfer function for uipolar output and a hyperbolic tan function for bipolar output.

7. Select a development system, such a s Neural Network Toolbox in MATLAB.

8. Select appropriate learning coefficients (η) and momentum factor(μ).

9. Select an acceptable training error ξ and a number of epochs. The training will stop whichever criterion was met earlier.

10. After the training is complete with all the patterns, test the network performance with some intermediate data points.

11. Finally, download the weights and implement the network by hardware or software.

8.9 Feed Back Signal Estimator for Vector Control

Neural network estimator is designed to estimate the feedback signals rather than conventional or digital estimator. A feed forward neural network based estimator estimates the rotor flux(ψr) ,unit vector($\cos\theta e, \sin\theta e$) and torque($T_e$) by solving the following equations:

$$\Psi_{dm}{}^{s} = \Psi_{ds}{}^{s} - i_{ds}{}^{s}L_{ls} \quad (8.84)$$

$$\Psi_{qm}{}^{s} = \Psi_{qs}{}^{s} - i_{qs}{}^{s}L_{ls} \quad (8.85)$$

$$\Psi_{dr}{}^{s} = L_r/L_m\, \Psi_{dm}{}^{s} - L_{lr}\, i_{ds}{}^{s} \quad (8.86)$$

$$\Psi_{qr}{}^{s} = L_r/L_m\, \Psi_{qm}{}^{s} - L_{lr}\, i_{qs}{}^{s} \quad (8.87)$$

$$\Psi_r = \sqrt{\Psi_{dr}^{s\,2} + \Psi_{qr}^{s\,2}} \quad (8.88)$$

$$\cos\theta_e = \Psi_{dr}{}^{s}/\Psi_r \quad (8.89)$$

$$\sin\theta_e = \Psi_{qr}{}^{s}/\Psi_r \quad (8.90)$$

$$T_e=(3P/4)*(\Psi_{dr}^{s}\, i_{ds}^{s} - \Psi_{qr}^{s}\, i_{ds}^{s}), \tag{8.91}$$

For the neural network input vectors(Ψ_{ds}^{s}, Ψ_{qs}^{s}, i_{ds}^{s}, i_{qs}^{s}) and target vectors (Ψ_r, $\cos\theta_e$, $\sin\theta_e$, T_e) are obtained from the induction motor model using neural network block This block calculate the required eight signals and send to the workspace

Neural network is of feedforward type, with three layers are used. In which four weight for input layer, twenty weights for hidden layer and four weights for output layer has been used. The input layer neurons have linear activation characteristics the hidden and output layers have a hyperbolic tan-type activation function to produce bipolar outputs Back propagation training method is used

Following is the neural network program.

```
%MATLAB code for ANN generation
net=newff(minmax(p),[4,20,4],{'purelin','tansig','tansig'},'trainscg')
net.trainParam.goal=1e-3;
net.trainParam.epochs=5000;net.trainParam.lr = 0.01;
[net,tr]=train(net,p,t);
a = sim(net,p),
```

Above given program meets the goal at 4846 epochs, 0.01 learning rate ,and the goal performance is 0.001, as shown in fig. 8.22

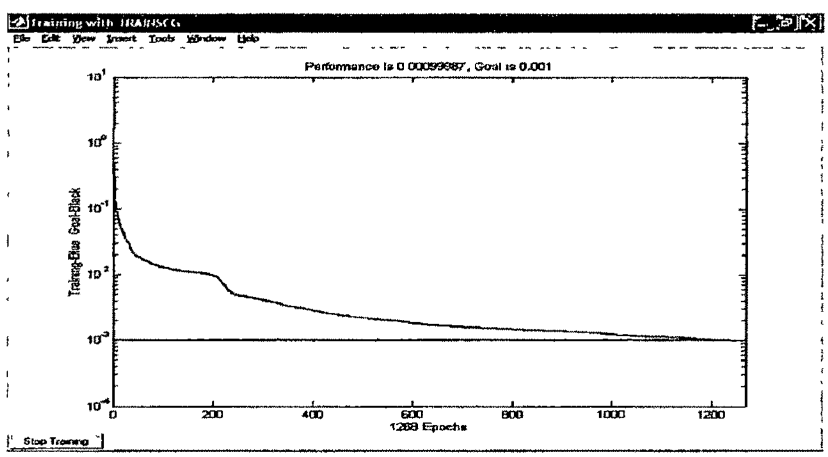


Fig 8.22 Training and goal versus Epochs

8.10 SPEED CONTORL OF INDUCTION MOTOR

A 3-Φ induction motor is practically a constant speed machine, more or less like a DC shunt motor. The speed regulation of an induction motor is usually less than 5% at full-load. However, DC shunt motors can made to run at any speed within wide limits, with good efficiency and speed regulation, merely by manipulating a simple field rheostat, the same is not possible with induction motors as the speed reduction is accompanied by a corresponding loss of efficiency and good speed regulation.

Different methods, by which speed control of induction motors is achieved, may be grouped as:

- **Control from stator side:**

1. By changing the applied voltage: This method, though the cheapest and the easiest, is rarely used because a large change in voltage is required for a relatively small change in speed. This large change in voltage will result in a large change in the flux density thereby seriously disturbing the magnetic conditions of the motor.

2. By changing the applied frequency: This method is also used very rarely. The synchronous speed of an induction motor is given by $N_s = 120f/P$. clearly, the synchronous speed (and hence the running speed) of an induction motor can be changed by changing the supply frequency f. However, this method could only be used in cases where the induction motor happens to be the only load on the generators, in which case, the supply frequency could be controlled by controlling the speed of the prime movers of the generators. But, here again the range over which the motor speed may be varied is limited by the economical speeds of the prime movers. This method has been used to some extent on electrically driven ships.

3. By changing the number of stator poles: This method is easily applicable to squirrel-cage motors because the squirrel-cage rotor adopts itself to any reasonable number of stator poles.

   This change of number of poles is achieved by having two or more entirely independent stator windings in the same slots. Each winding gives a different number of poles and hence different synchronous speed. For example, a 36-slot stator may have two 3- windings, one with 4 poles and the other with 6-poles. With a supply frequency of 50 Hz, 4-pole winding will give $N_s = 120 * 50/4 = 1500$ r.p.m. and the 6- pole winding will give $N_s = 120*50/6 = 1000$ r.p.m. Motors with four independent stator winding are also in use and they give four different synchronous (and hence running) speeds. Of course, one winding is used at a time, the others being entirely disconnected. This method has been used for elevator motors, traction motors and also for small motors driving machine tools.

   Speed in the ration of 2:1 can be produced by a single winding if wound on the consequent, pole principle. In that case, each of the two stator windings can be connected by a simple switch to give two speeds, each, which means four speeds in all. For example, one stator winding may give 4 or 8 poles and the other 6 or 12 poles. For a supply frequency of 50 Hz, the four speeds will be 1500, 750, 1000 and 500 r.p.m. another combination, commonly used, is to group 2 and 4 pole winding with a 6 and 12 winding, which gives four synchronous speeds of 300, 1500, 1000, 500 r.p.m.

- **Control from rotor side:**

1. Rotor rheostat control: In this method, which is applicable to slip-ring motors alone, the motor speed is reduced by introducing an external resistance in the rotor circuit. For the purpose, the rotor starter may be used, provided it is continuously rated. This method is, in face, similar to the armature rheostat control method of DC shunt motors. For induction motor, near synchronous peed (i.e. for very small slip value), T Q s/R_2 it is obvious that for a given torque, slip can be increased i.e. speed can be decreased by increasing the rotor resistance R_2. it is used where speed changes are needed for short periods only.

2. By operating two motors in concatenation or cascade(Fig 8.23)

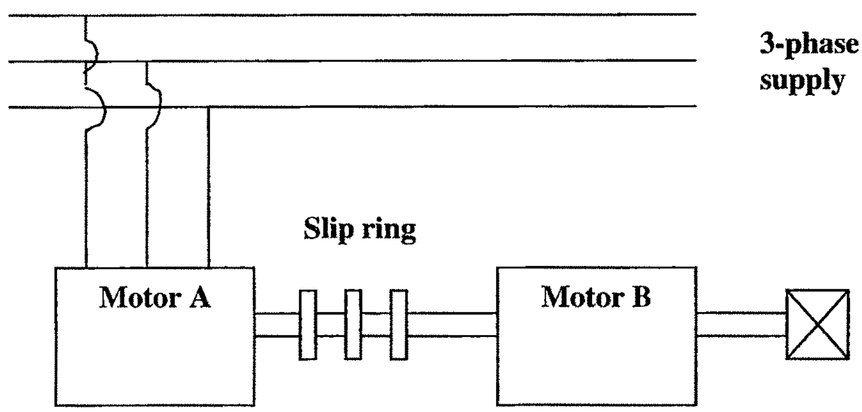


Fig 8.23: Cascade operation.

In this method, two motors are used and are ordinarily mounted on the same shaft, so that both run at the same speed (or else they may be geared together). The stator winding of the main motor A is connected to the mains in the usual way, while that of the auxiliary motor B is fed from the rotor circuit of motor A. for satisfactory operation, the main motor A should be phase-wound i.e. of slop-ring type with stator to rotor winding ratio of 1:1, so that, in addition to concatenation, each motor may be run from the supply mains separately. There are at least three ways in which the combination may be run.

- Main motor A may be run separately from the supply. In that case, the synchronous speed is $N_{sa} = 120f/P_a$ where P_a = Number of stator poles of motor A.
- Auxiliary motor B may be run separately from mains (with motor A being disconnected). In that case, synchronous speed is $N_{sb}=120f/P_b$ where P_b = number of stator poles of motor B.
- The combination may be connected in cumulative cascade i.e. in such a way that the phase rotation of the stator fields of both motors is in the same direction. The synchronous speed of the cascaded set, in this

case, is $N_{sc} = 120f/(P_a + P_b)$.

3. **By injection and e.m.f. in the rotor circuit:** In this method, the speed of an induction motor is controlled by injecting a voltage in the rotor circuit, it being of course, necessary for the injected voltage to have the same frequency as the slip frequency. There is, however, no restriction as to the phase of the injected e.m.f. By changing the phase of the injected e.m.f.(rotor resistance), speed can be controlled.

The *induction motor drives* can be controlled by scalar control, vector or field-oriented control, direct torque and flux control, adaptive control.

1. **Scalar control:** Scalar control, as the name indicates, is due ot magnitude variation of the control variables only, and 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. Scalar control is in contrast to vector or field-oriented control, where both the magnitude and phase alignment of vector variables are controlled.
2. **Vector or field-oriented control:** The problem of scalar control can be solved by vector or filed-oriented control. The invention of vector control in the beginning of 1970s, and the demonstration that an induction motor can be controlled like a separately excited dc motor; brought a renaissance in the high-performance control of ac drives. Because of dc machine like performance, vector control is also known as decoupling, orthogonal, or transvector control.
3. **Sensorless vector control:** Sensorless vector control of an induction motor drive essentially means vector control without any speed sensor. An incremental shaft-mounted speed encoder, usually an optical type, is required for close loop speed or position control in both vector and scalar controlled drives. A speed signal is also required in indirect vector control in the whole speed range, and in director vector control for the low-speed range, including the zero speed start-up operation. A speed encoder is undesirable in a drive because it adds cost and reliability problems, besides the need for a shaft extension and mounting arrangement. It is possible to estimate the speed signal form machine terminal voltages and currents with the help of a DSP. However, the estimation is normally complex and heavily dependent on machine parameters. Although sensorless vector controlled drives are commercially available at this time, the parameter variation problem, particularly near zero speed, imposes a challenge in the accuracy of speed estimation.

4. **Adaptive Control:** A linear control system with invariant plant parameters can be designed easily with the classic design techniques. Ideally, a vector-controlled ac drive can be considered as linear, like a dc drive system. However, in industrial applications, the electrical and mechanical parameters of the drive hardly remain constant. Besides, there is a load

torque disturbance effect. The effect of parameter variation can be compensated to some extent by a high-gain negative feedback loop. But, excessive gain may cause an underdamping or instability problem in extreme cases. These problems require adaptation of the controller in real time, depending on the plant parameter variation and load torque disturbance, so that the system response is not affected. Adaptive control techniques can be generally classified as

* Self-tuning control
* MARC
* Sliding mode or variable structure control.
* Expert system control.
* Fuzzy control.
* Neural control

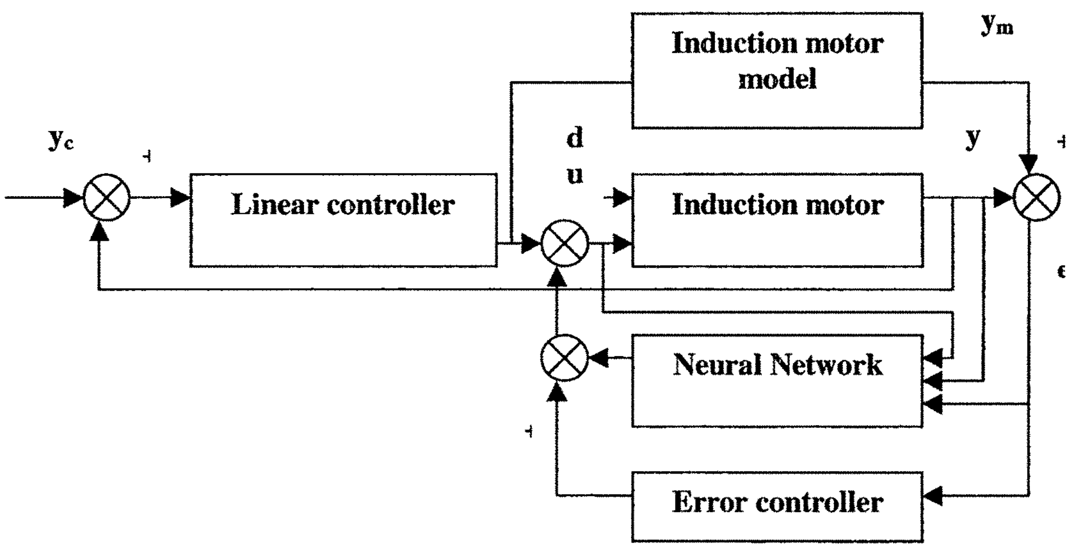


Fig 8.24: Block diagram for speed control of induction motor.

In this approach the difference between the speeds of the induction motor and model is used by the neural network for estimation and adjustment of control signal applied to the induction motor.

8.10.1 Design of Graphical User Interface (GUI):

A graphical user interface (GUI) is a user interface built with graphical objects; the components of the GUI; such as buttons, text fields, sliders, and menus. If the GUI is designed well-designed, it should be intuitively obvious to the user how its components function. Applications that provide GUIs are generally easier to learn and use since the person using the application does not need to know what commands are available or how they work. The action that results from a particular user action can be made clear by design of the interface.

MATLAB implements GUIs as figure windows containing various styles of uicontrol objects. We must program each object to perform the intended action when activated by user of the GUI. All of these tasks are simplified by (GUIDE) MATLAB's Graphical User Interface Development Environment.

- **Implementation Of GUI:** Creating a GUI involves following basic tasks: Laying out and programming components, saving and running GUI.

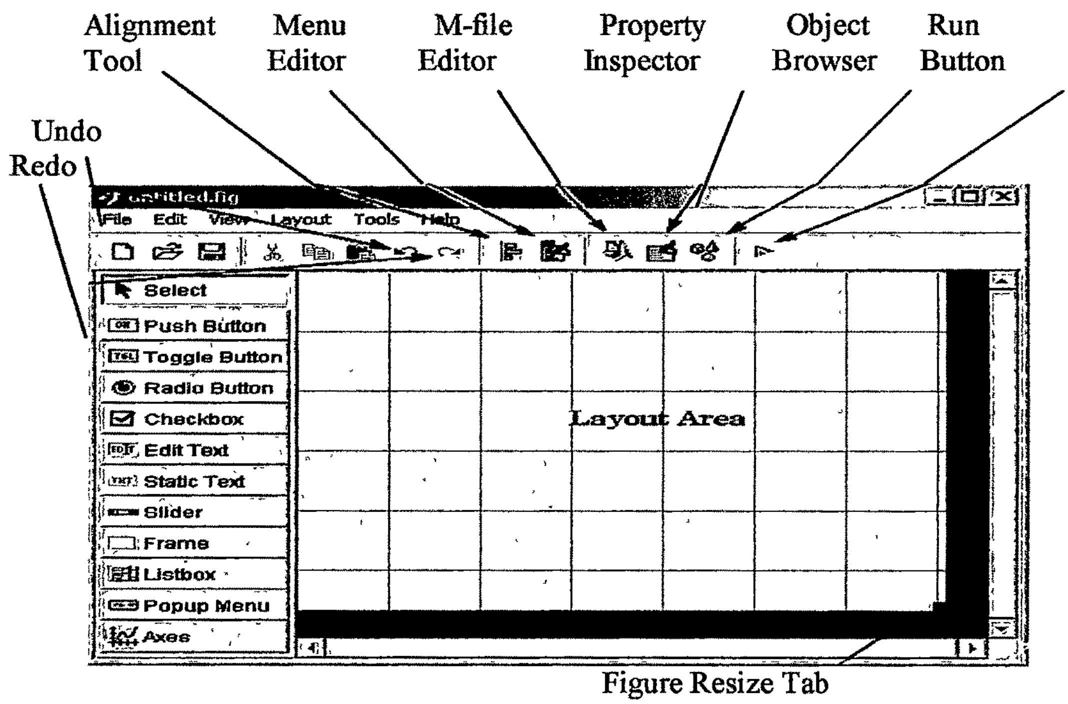


Fig 8.24: Layout Editor.

GUIDE primarily is a set of layout tools (Fig 8.24) However, GUIDE also generates an M-file that contains code to handle the initialization and launching of the GUI. This M-file provides a framework for the implementation of the callbacks, the functions that execute when users activate components in the GUI. GUIDE stores GUIs in two files, which are generated the first time we save or run the GUI:

1. *FIG-file:* A file with extension .fig that contains a complete description of the GUI figure layout and the components of the GUI: push buttons, menus, axes, and so on. When we make changes to the GUI layout in the Layout Editor, our changes are saved in the FIG-file.

2. *M-file:* A file with extension .m that contains the code that controls the GUI, including the callbacks for its components. This file is referred to as the GUI M-file. When we first run or save a GUI from the Layout Editor, GUIDE generates the GUI M-file with blank stubs for each of the callbacks. We can than program the callbacks using the M-file editor.

Fig 8.25 illustrates the parts of GUI implementation.

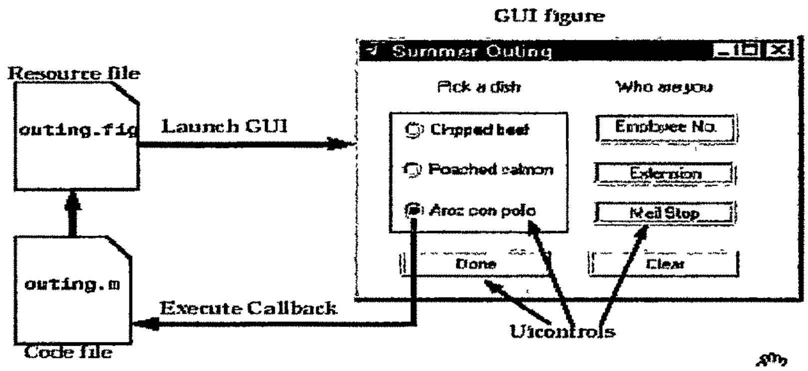
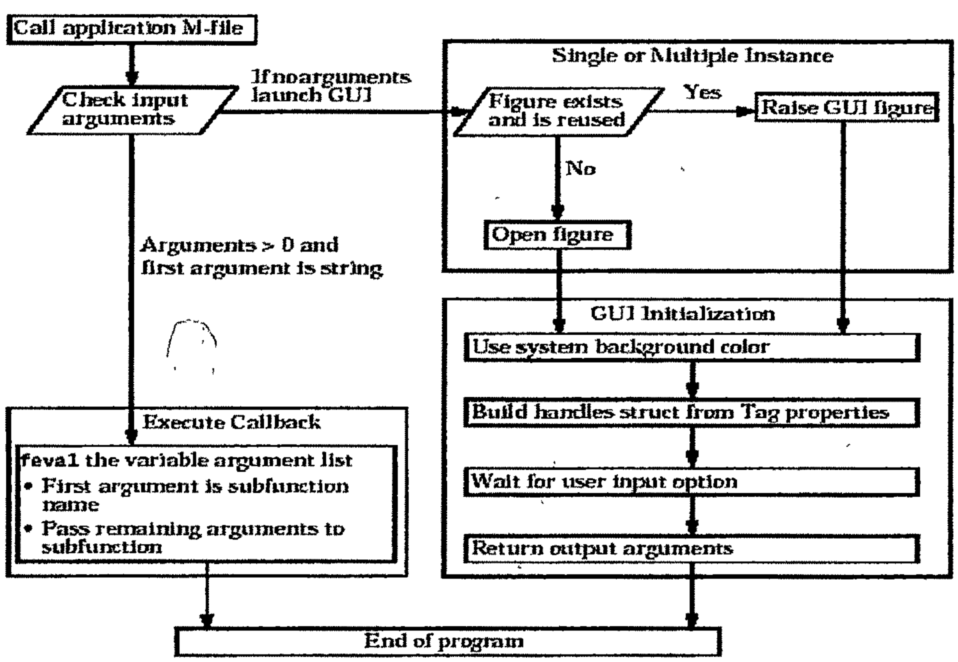


Fig 8.25: Parts of GUI implementation.

- **Programming the GUI components:** The GUI M-file generated by GUIDE controls the GUI and determines how it responds to a user's actions, such as pressing a push button or selecting a menu item. The M-file contains all the code needed to run the GUI, including the callbacks for the GUI components. While GUIDE generates the framework for this M-file, we must program the callbacks to perform the functions we want them to. When we run a GUI, the M-file creates a handles structure that contains all the data for GUI objects, such as controls, menus, and axes. The handles structure is passed as an input to each callback. We can use the handles structure to Share data between callbacks and access the GUI data. We can add code to the Opening function; executes before the GUI becomes visible to the user, Output function; outputs data to the command line, if necessary and in Callbacks; execute each time the user activates the corresponding component of the GUI.



9

Fig 8.26: M-file Execution path.

After writing the callbacks, we can run the GUI by selecting Run from the Tools menu of clicking the Run button on the GUIDE toolbar.

A GUI can be created that sets the parameters of SIMULINK model. The SIMULINK model of the application must be open before running its GUI. If it is closed then GUI reopens it whenever it requires the model to execute. In addition, giving the facilities of obtaining simulation and saving the results and also plotting the results, the GUI can run the simulation and plot the results.

8.10.2 GUI *for Robust Adaptive observer based ANN estimator* Robust Controller for Induction motor Drive

The optimal control problem requires the state estimate at every instant of time. The state vector of the system can be constructed from its inputs and available outputs employing observer-estimator provided, system is completely observable and then the observer system is completely controllable. Even if observer output is noisy then also the states can be estimated through robust adaptive optimization techniques. (H^∞ / H^2 norm). Paper describes design and simulation of a robust adaptive observer.

The robust adaptive observer is used for training the neural network to generate feed back signal estimate. The generated NN block is used for closed loop current control of induction motor within permissible range. GUI design for speed control of induction motor is implemented, which allows the user to change the parameter of induction motor, to select Robust or Non-robust (Kalman Filtering)operation, to initialize and run simulation model to generate training pairs. Generated ANN block for new parameters can be used as a library block for the SIMULINK models in other applications.

The SIMULINK model developed in fig 8.14 is used. Table 8.3 gives a list of MATLAB .m and SIMULINK .mdl files to be linked with GUI components.

FILE NAME	STORED DATA	PURPOSE
motorparameter m	Performance parameters of motor like R_s, R_r, L_{ls}, L_{lr}, P, J, X_{lr}, X_{ls}, w_b, T_r, L_r, A, B, C, D. System Matrices of Induction motor model.	On execution, all the parameters will be stored at workspace.
rhzeros m	Program:H^∞ controller with zero padding.	On execution optimal value of gamma, xinf, yinf, K_c, K_e, motor model in augmented space are obtained and stored at workspace.
dhzeros.mdl	Simulink model for H^∞ controller	Running Simulation robust adaptive observer is simulated using the stored data in workspace.

Table 8.3: MATLAB/ SIMULINK File Details

Fig 8.27 depicts the GUI for our application stored as figure file ***ann_esimator.fig*** as well as MATLAB m-file ***ann_estimator.m***

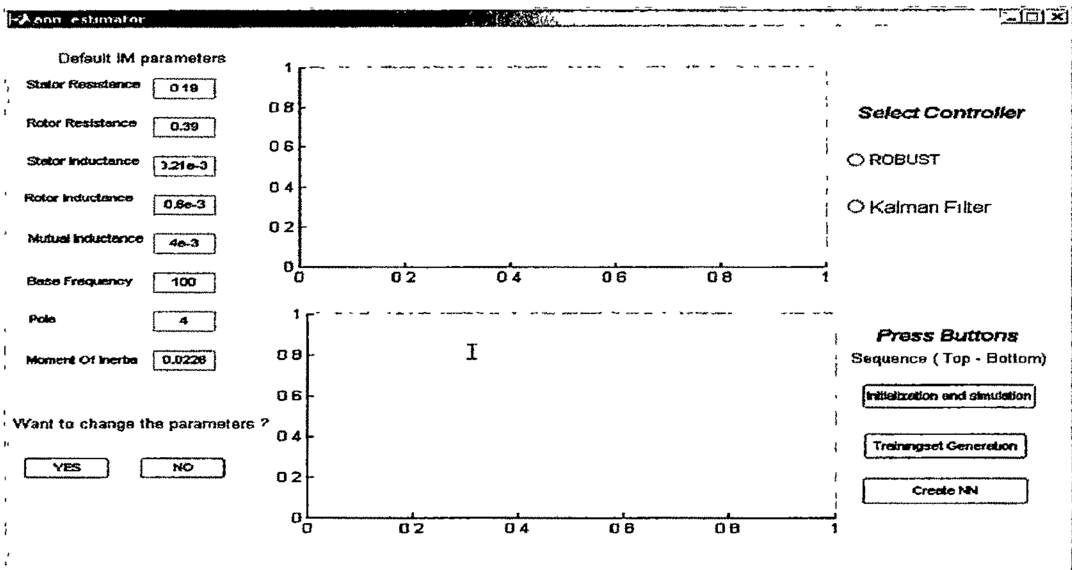


Fig 8.27: Graphical User Interface: Generation of ANN Estimator

The operating procedure is as follows

I. Open the interface figure file ***ann_estimator.fig*** through GUIDE or run the file ann_esimator.*m* in command window of MATLAB Screen of Fig 8 27 will appear.

II. Default parameters of induction motor are specified. User may change the motor parameter by pressing *YES* button otherwise *NO* button is pressed to proceed.

III. User may select Robust or Non-robust operation

IV. On pressing the initialization and simulation button the model simulink induction motor *model motorcontrol.mdl* will be visible on screen and executed.

V On Pressing the *triggering set generation* button the training set for neural Network will be generated and Fig 18.28 will appear.

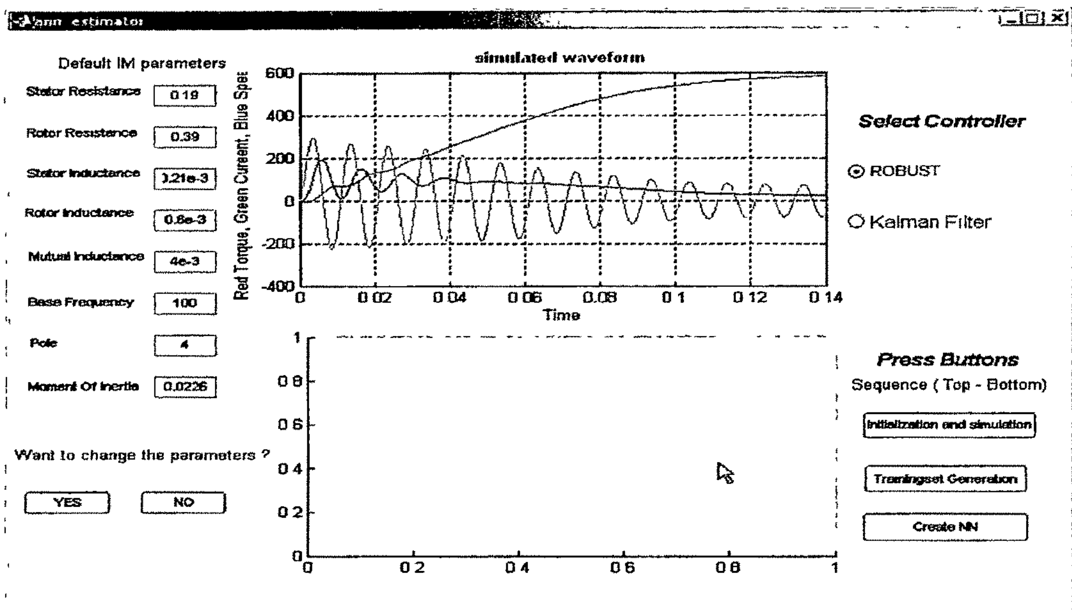


Fig 8.28: Graphical User Interface: Generation of ANN Training Set

VI. Pressing *create NN* button will train neural network and NN block (Fig 8.29) is Generated

It is a feed forward ANN with three layers, has 4 nodes in the input Layer, 20 nodes in the hidden layer and 4 nodes in the output layer is used The input layer neurons have linear activation characteristics while the hidden and output layers have a hyperbolic tan-type activation function to produce bipolar outputs Training method used is Back propagation

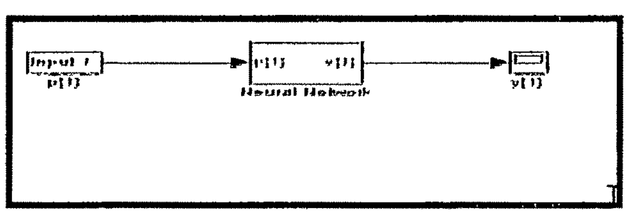


Fig 8.29: Generated Block: ANN Estimator

VII. ANN block generated is used as a library symbol for operation (Fig 8.30) The results are similar to those obtained in previous sections

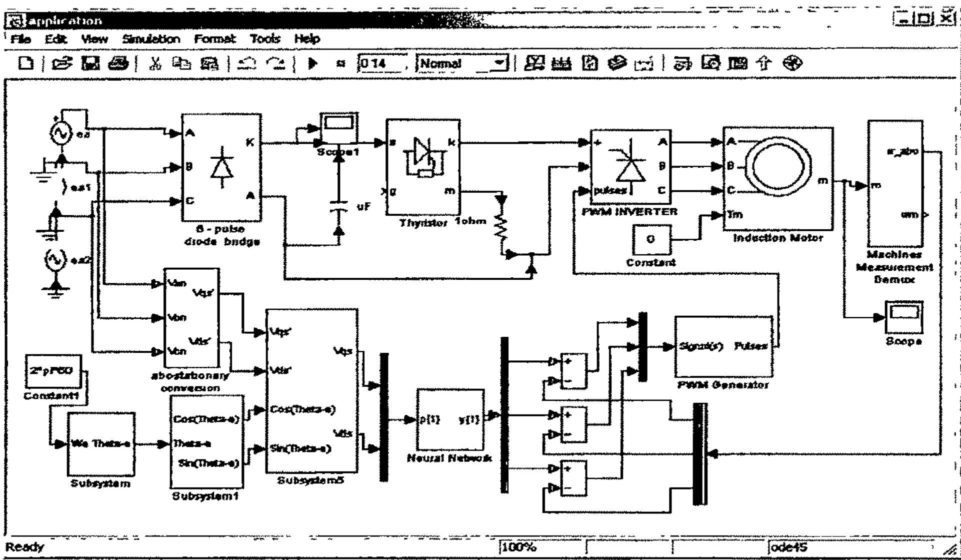


Fig 8.30: SIMULINK Model: ANN based Speed Control of Induction Motor

The design and simulation of a robust adaptive observer is described The robust adaptive observer is used for training the neural network to generate feed back signal estimate The generated NN block is used for closed loop current control of induction motor within permissible range. GUI design for speed control of induction motor is implemented, which allows the user to change the parameter of induction motor, to select Robust or Non-robust (Kalman Filtering) operation, to initialize and run simulation model to generate training pairs. Generated ANN block for new parameters can be used as a library block for the SIMULINK models in a speed control application