CHAPTER 5

CONTROL SCHEME AND CONTROLLER DESIGN FOR INDUCTION MOTOR DRIVES

5.1 Introduction

Induction motor drives have been and are the workhorses in the industry for variable speed applications in a wide power range that covers from fractional horsepower to multi-megawatts. These applications include pumps and fans, paper and textile mills, subway and locomotive propulsions, electric and hybrid vehicles, machine tools and robotics, wind power generation systems, etc. In this chapter the control of induction motor drives for variable speed applications is explained. The control schemes available for the induction motor drives are the scalar control, vector or field oriented control, direct torque and flux control and adaptive control. In this chapter special emphasis is on vector control of induction motor, though some aspects of scalar control for induction machine is studied.

5.2 Scalar Control Of Induction Machine

Scalar control as the name indicates is due to magnitude variation of the control variables only, and disregards any coupling effect in the machine. For example the voltage of the machine can be controlled to control the flux, and the frequency or slip can be controlled to control torque.

However, flux and torque are also functions of frequency and voltage respectively. In scalar control both the magnitude and the phase alignment of vector variables are controlled. Scalar controlled drives give somewhat inferior performance than the other control schemes but they are the easy to implement. In the following sections scalar control techniques with voltage-fed inverters are discussed.

5.2.1 Open Loop Volts/Hz Control [15]

The open loop Volts/Hz control of an induction motor is far the most popular method of speed control because of its simplicity and these types of motors are widely used in industry. Traditionally, induction motors have been used with open loop 60Hz power supplies for constant speed applications. For adjustable speed applications, frequency control is natural. However, voltage is required to be proportional to frequency so that the stator flux ($\psi_s = \frac{V_s}{\omega_e}$) remains constant if the stator resistance is neglected. Figure 5.1 shows the block diagram of the open loop Volts/Hz control method. The power circuit consists of a diode rectifier with a single or three-phase ac supply, filter and PWM voltage-fed inverter. Ideally no feedback signals are required for this control scheme. The frequency ω_e^* is the primary control variable because it is approximately equal to the rotor speed ω_r neglecting the slip speed, as it is very small (ideally).

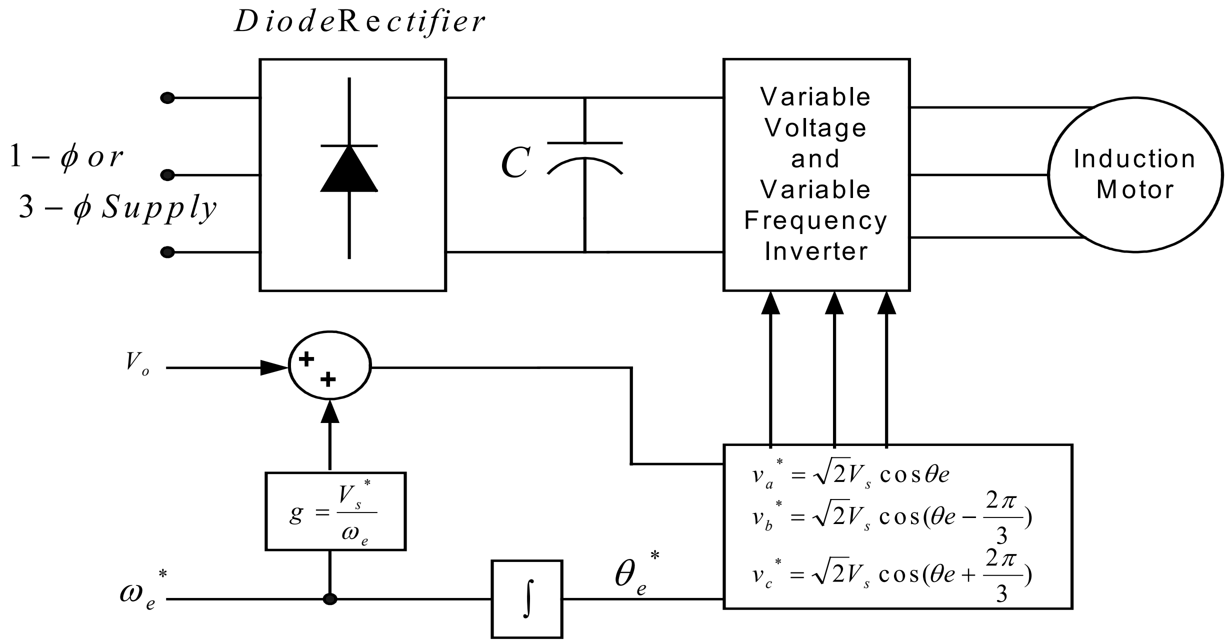

Figure 5.1: Block diagram of the open loop Volts/Hz control for an induction motor.

The phase voltage V_s^* command is directly generated from the frequency command by the gain factor 'g', as shown, so that the flux ψ_s remains constant. If the stator resistance and the leakage inductance of the machine are neglected then the flux will also correspond to the air gap flux ψ_m or rotor flux ψ_r. As the frequency becomes small at low speed, the

stator resistance tends to absorb the major amount of the stator voltage, thus weakening the flux.

The boost voltage V_o is added so that the rated flux and corresponding full torque become available down to zero speed. The effect of this boost voltage V_o at high frequencies is small. The ω_e^* signal is integrated to generate the angle signal θ_e^*, and the corresponding sinusoidal phase voltages (v_a^*, v_b^*, v_c^*) signals are generated as

$$v_a^* = \sqrt{2} V_s \cos\theta_e \tag{5.1}$$

$$v_b^* = \sqrt{2} V_s \cos(\theta_e - \frac{2\pi}{3}) \tag{5.2}$$

$$v_c^* = \sqrt{2} V_s \cos(\theta_e + \frac{2\pi}{3}) \tag{5.3}$$

The PWM converter is merged with the inverter block. Some problems encountered in the operation of this open loop drive are the following [15]:

(1) The speed of the motor cannot be controlled precisely, because the rotor speed will be slightly less than the synchronous speed and that in this scheme the stator frequency and hence the synchronous speed is the only control variable.

(2) The slip speed, being the difference between the synchronous speed and the electrical rotor speed, cannot be maintained, as the rotor speed is not measured in this scheme. This can lead to operation in the unstable region of the torque-speed characteristics.

(3) The effect of the above can make the stator currents exceed the rated current by a large amount thus endangering the inverter-converter combination.

These problems are to an extent overcome by having an outer loop in the induction motor drive, in which the actual rotor speed is compared with its commanded value, and the error is processed through a controller usually a PI controller and a limiter is used to obtain the slip-speed command.

The limiter ensures that the slip-speed command is within the maximum allowable slip-speed of the induction motor. The slip-speed command is added to electrical rotor speed to obtain the stator frequency command. Thereafter the stator frequency command is processed as in an open loop drive. In the closed loop induction motor drive the limits on the slip speed, boost voltage and reference speed are externally adjustable variables. The external adjustment allows the tuning and matching of the induction motor to the converter and inverter and the tailoring of its characteristics to match the load requirements.

5.3 Vector Control Of Induction Motor

Scalar control is simple to implement, but the inherent coupling effect (that is both the flux and the torque are functions of voltage or current and frequency) gives sluggish response and the system is prone to instability because of a high order system effect. If the torque is increased by incrementing the slip or frequency the flux tends to decrease and this flux variation is very slow. The flux decrease is then compensated by the flux control loop, which has a large time constant. This temporary dipping of flux reduces the torque sensitivity with slip and lengthens the response time. The variations in the flux linkages have to be controlled by the magnitude and frequency of the stator and rotor phase currents and their instantaneous phases. Normal scalar control of induction machine aims at controlling the magnitude and frequency of the currents or voltages but not their phase angles.

Separately excited dc motor drives are simple in control because they independently control flux, which, when maintained constant, contributes to an independent control of torque. This is made possible with separate control of field and armature currents, which, in turn control the field flux and the torque independently. Moreover the dc motor control requires only the control of the field or armature current magnitudes, providing simplicity not possible with an ac machine. In contrast, the induction motor drive requires a coordinated control of stator current magnitudes, frequencies and phase magnitude making it a complex control. As with the dc motor drives, independent control of flux and the torque is possible in ac drives.

The stator current phasor can be resolved along the rotor flux linkages and the component along the rotor flux linkages is the field producing current, but this requires the position of the rotor flux linkages at every instant. If this is available then the control of the ac machines is very similar to that of the dc drives. The requirement of phase, frequency and magnitude control of the currents and hence the flux phasor is made possible by inverter control. The control is achieved in field coordinates and hence it is often called the field-oriented control or the vector control, because it relates to the phasor control of the flux linkages.

5.3.1 Principle Of Vector Control [18, 19, 20]

The vector control of induction machine is explained by assuming that the position of the rotor flux linkages phasor λ_r is known. The phasor diagram of the vector control is as shown in figure 5.2. λ_r is at θ_f from the stationary reference, θ_f is referred to as field angle.

The three-stator currents can be transformed into q and d axed currents in

the synchronous reference frames by using the transformation,

$$\begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos\theta_f & \cos(\theta_f - \frac{2\pi}{3}) & \cos(\theta_f + \frac{2\pi}{3}) \\ \sin\theta_f & \sin(\theta_f - \frac{2\pi}{3}) & \sin(\theta_f + \frac{2\pi}{3}) \end{bmatrix} \begin{bmatrix} i_{as} \\ i_{bs} \\ i_{cs} \end{bmatrix} \quad (5.4)$$

from which the stator current phasor i_s is derived as

$$|i_s| = \sqrt{(i_{qs})^2 + (i_{ds})^2} \quad (5.5)$$

and the stator phase angle is

$$\theta_s = \tan^{-1}(\frac{i_{qs}}{i_{ds}}) \quad (5.6)$$

where i_{qs} and i_{ds} are the ‘q’ and the ‘d’ axes currents in the synchronous reference frame that are obtained by projecting the stator current phasor on the ‘q’ and ‘d’ axes respectively. The current phasor magnitude remains the same regardless of the reference frame chosen as is clear from Figure 5.2.

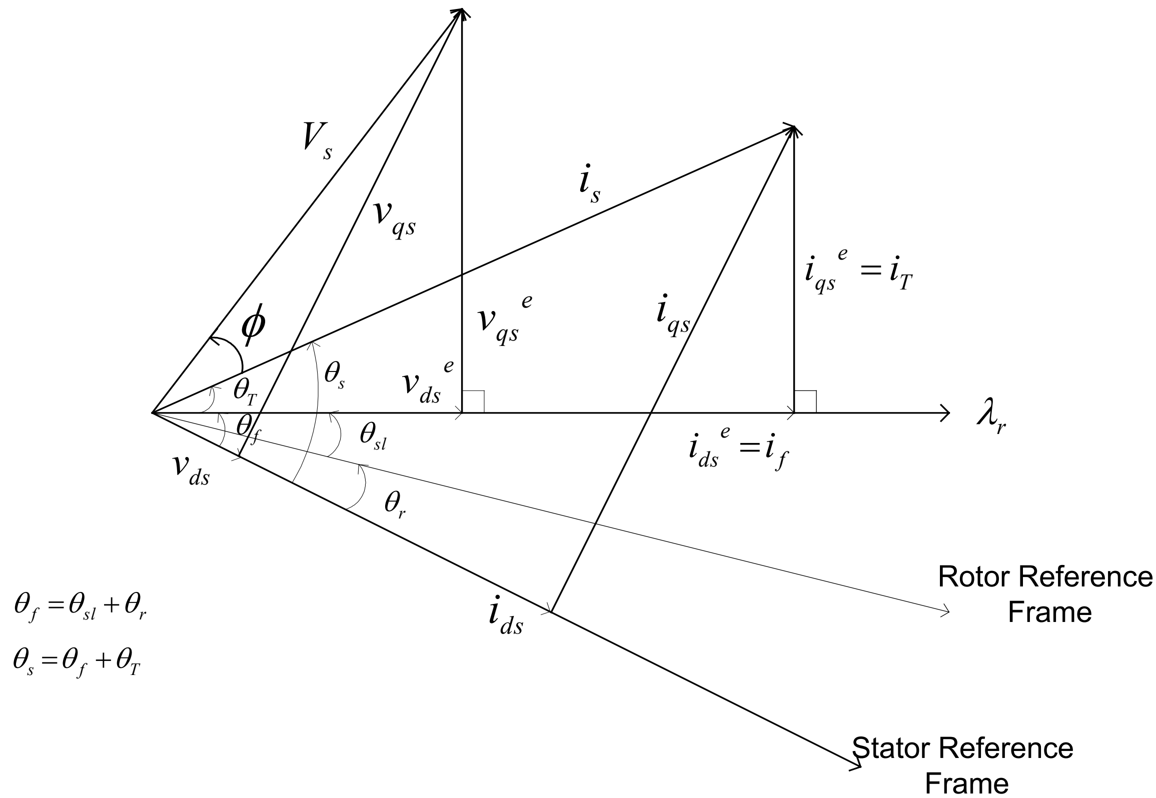

Figure 5.2: Phasor diagram of the vector controller.

The stator current i_s produces the rotor flux λ_r and the torque T_e. The component of current producing the rotor flux has to be in phase with λ_r. Therefore, resolving the stator current phasor along λ_r reveals that the component i_f is the field-producing component as shown in Figure 5.2. The perpendicular component is hence the torque-producing component i_T. Thus,

$$\lambda_r \propto i_f \tag{5.7}$$

$$T_e \propto \lambda_r i_T \propto i_f i_T \ . \tag{5.8}$$

The components i_f and i_T are only dc components in steady state, because the relative speed with respect to that of the rotor field is zero. Orientation of λ_r amounts to considering the synchronous reference frame and hence the flux and the torque-producing components of currents are dc quantities, and this makes them to be used as control variables. Till now it was assumed that the rotor flux position is actually available, but it has to be obtained at every instant. This field angle can be written as,

$$\theta_f = \theta_{sl} + \theta_r \tag{5.9}$$

where θ_r is the rotor position and θ_{sl} is the slip angle. In terms of the speeds and time, the field angle can be written as

$$\theta_f = \int (\omega_r + \omega_{sl}) dt = \int \omega_s \, dt \tag{5.10}$$

Vector control schemes thus depend upon how the instantaneous rotor flux position is obtained and are classified as direct and the indirect vector control schemes.

In the direct vector control the field angle is calculated by using terminal voltages and currents or Hall sensors or flux sensing windings. The rotor flux position can also be obtained by using rotor position measurement and partial estimation with only machine parameters but not any other variables, such as voltages and currents. Using this field angle is called the indirect vector control.

In this thesis the indirect vector control for the induction motor is employed so further discussions on the vector control scheme will be emphasized on indirect vector control.

5.3.2 Derivation Of Indirect Vector Control For Induction Motor

In the derivation for the indirect vector control scheme a current source inverter is assumed, in which case the stator phase currents serve as inputs, hence the stator dynamics can be neglected. The dynamic equations of the induction motor in the synchronous reference frame for the rotor taking flux as state variable is given as,

$$r_r i_{qr} + p\lambda_{qr} + \omega_{sl}\lambda_{dr} = 0 \tag{5.11}$$

$$r_r i_{dr} + p\lambda_{dr} - \omega_{sl}\lambda_{qr} = 0 \tag{5.12}$$

where

$$\omega_{sl} = \omega_s - \omega_r \tag{5.13}$$

$$\lambda_{qr} = L_r i_{qr} + L_m i_{qs} \tag{5.14}$$

$$\lambda_{dr} = L_r i_{dr} + L_m i_{ds} \quad . \tag{5.15}$$

The definition for the different symbols was given in chapter 3 and so is not repeated here. The resultant rotor flux linkage, λ_r also known as the rotor flux linkages phasor is assumed to be on the direct axis to achieve field orientation. This alignment reduces the number of variables to deal.

The alignment of the d-axis with rotor flux phasor yields

$$\lambda_r = \lambda_{dr} \tag{5.16}$$

$$\lambda_{qr} = 0 \tag{5.17}$$

$$p\lambda_{qr} = 0 \tag{5.18}$$

Substituting Equations 5.15 to 5.17 in Equations 5.11 and 5.12 causes the new rotor equations to be,

$$r_r i_{qr} + \omega_{sl} \lambda_r = 0 \tag{5.19}$$

$$r_r i_{dr} + p\lambda_r = 0 \tag{5.20}$$

Thus from Equations 5.14 and 5.15 the rotor currents are derived as in Equations 5.21 and 5.22.

$$i_{qr} = \frac{-L_m}{L_r} i_{qs} \tag{5.21}$$

$$i_{dr} = \frac{\lambda_r - L_m i_{ds}}{L_r} \ . \tag{5.22}$$

substituting for d and q axes rotor currents from equations 5.21 and 5.22 into Equations 5.19 and 5.20 the following equations are obtained.

$$i_f = \frac{1}{L_m}[1 + T_r p]\lambda_r \tag{5.23}$$

From Equation (5.19),

$$\omega_{sl} = -\frac{r_r i_{qr}}{\lambda_r} = \frac{L_m}{T_r} \frac{i_T}{\lambda_r} \tag{5.24}$$

where

$$i_T = i_{qs} \tag{5.25}$$

$$i_f = i_{ds} \tag{5.26}$$

$$T_r = \frac{L_r}{r_r} \tag{5.27}$$

$$K_{it} = \frac{4}{3P} \ . \tag{5.28}$$

Equation 5.23 resembles the field equation in a separately excited dc machine whose time constant is usually in the order of seconds. Substituting the rotor currents, the torque expression can be obtained as

$$T_e = \frac{3P}{4}\frac{L_m}{L_r}(\lambda_{dr} i_{qs} - \lambda_{qr} i_{ds}) = \frac{3P}{4}\frac{L_m}{L_r}(\lambda_{dr} i_{qs}) = K_{te}(\lambda_r i_{qs}) \tag{5.29}$$

From Equation 5.29 it can be observed that torque is proportional to the product of the rotor flux linkages and the stator q-axis current. This resembles the air gap torque expression of the dc motor, which is proportional to the product of the field flux linkages and the armature current. If the rotor flux linkage is maintained constant then the torque is simply proportional to the torque-producing component of the stator current as in the case of the separately excited dc motor. Similar to the dc machine time constant, which is of the order of few milliseconds, the time constant of the torque current is also of the same order.

Equations 5.23 and 5.29 complete the transformation of the induction machine into an equivalent separately excited dc motor from control point of view. The stator current phasor is the sum of the 'd' and the 'q' axes stator currents in any reference frame given as

$$i_s = \sqrt{(i_{qs})^2 + (i_{ds})^2} \tag{5.30}$$

and the 'dq' axes to abc phase current relationship is obtained from Equation 5.4.

5.3.3. Implementation Of Indirect Vector Control

The indirect vector controller [16] takes only the speed from the machine while all other parameters are estimated. The implementation of the indirect vector control is as shown in Figure 5.3. The torque command is generated as a function of the speed error signal, generally processed through a PI controller. The flux command for a simple drive strategy is made to be a function of speed, defined by

$$\begin{aligned} \lambda_r^* &= \lambda_b & 0 \le |\omega_r| \le \omega_{rated} \\ &= \frac{\omega_b}{|\omega_r|}\lambda_b & \omega_b \le |\omega_r| \le \omega_{r(\max)} \end{aligned} \tag{5.31}$$

where $\lambda_b \, and \, \omega_r$ are the rated or the base rotor flux linkages and rotor speed respectively. The flux is kept at rated value up to rated speed; above that the flux is weakened to maintain the power output at a constant value. By doing the manipulations as shown in Figure 5.3 the three-phase stator current commands are generated.

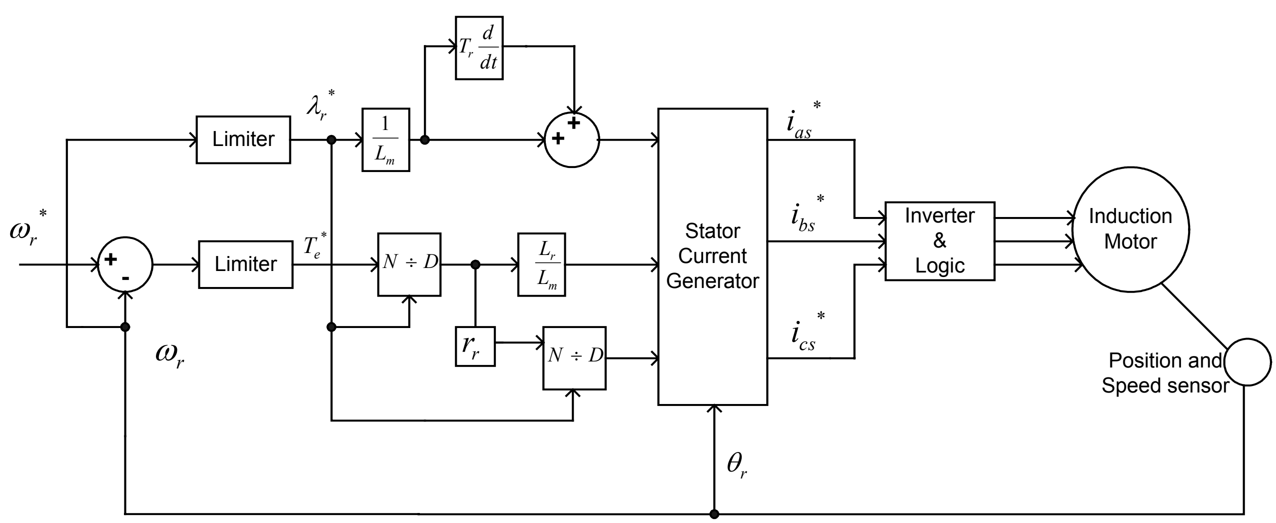

Figure 5.3: The implementation of the indirect vector control.

These current commands are simplified through a power amplifier, which can be any standard converter-inverter arrangement. The rotor position θ_r is measured with an encoder and converted into necessary digital signal for feedback.

In the above discussion the inverter assumed is a current source to simplify explanation but in this thesis a voltage source inverter is used so the current generated are converted into voltages and are given as the commands for the voltage source inverter.

5.4 Controller Design For Induction Motor

The control objective is to regulate the actual quantity with the reference command. As is discussed above the speed and the flux are controlled to get the reference for the currents, which can be regulated to synthesize the modulation signals for the inverter. Since the speed and flux are dc quantities to regulate these quantities normal PI, PD or PID controller can be used, but usually a PI controller achieves the best performance. Any quantity of the system can be set as the output of the controller provided a relation as to how the controlled quantity effects the variable taken as the output can be given, but this is usually a tough job.

A linearization technique is explained below which helps in the controller design and to decide what should be the output of the controller. After deciding the type of controller, a way to determine the parameters of the controller has to be set forth.

5.4.1. Feedback linearization Control

This control scheme is a type of non-linear control scheme whose design is based on exact linearization. The design technique consists of two steps as described below [21].

1. A nonlinear compensation, which cancels the nonlinearities included in the system, is implemented as an inner feedback loop.
2. A controller, which ensures stability and some predefined performance, is designed based on the conventional theory; this linear controller is implemented as an outer feedback loop. Consider the third order system

$$px_1 = \sin x_2 + (x_2 + 1)x_3 \tag{5.32}$$

where 'p' is the differential operator $p = \frac{d}{dt}$.

$$px_2 = x_1^5 + x_3 \tag{5.33}$$

$$px_3 = x_1^2 + u \tag{5.34}$$

$$y = x_1 \tag{5.35}$$

To generate a direct relationship between the output y and the input u, let us differentiate the output 'y'

$$py = px_1 = \sin x_2 + (x_2 + 1)x_3 \tag{5.36}$$

since py is still not directly related to the input u, the differentiation is carried out once again to obtain,

$$p^2 y = (x_2 + 1)u + f_1(x_1, x_2, x_3) \tag{5.37}$$

where $f_1(x_1, x_2, x_3) = (x_1^5 + x_3)(\cos x_2 + x_3) + (x_2 + 1)x_1^2$.

Now an explicit relation exists between y and u. If the control input is chosen to be in the form

$$u = \frac{1}{x_2 + 1}(v - f_1) . \tag{5.38}$$

where v is a new input to be determined, the nonlinearity in the above equation is canceled and a simple linear double integration relationship between the output and the new input v is obtained.

$$p^2 y = v \ . \tag{5.39}$$

The design of ac controller for this doubly-integrator relationship is simple, because of the availability of linear control techniques. For instance, let us define the error as,

$$e = y(t) - y_d(t) \tag{5.40}$$

where $y_d(t)$ is the desired output. Choosing the new input v as

$$v = p^2 y - k_1 e - k_2 pe \tag{5.41}$$

with k_1 and k_2 being positive constants , the error of the closed loop system is given by

$$p^2 e + k_2 pe + k_1 e \ . \tag{5.42}$$

which represents an exponentially stable error dynamics.

If initially e (0)= pe(0) = 0, then e(t) =0, and a perfect control is achieved , otherwise, e(t) converges to zero exponentially.

The feedback linearization technique is used in this thesis to design the controller for induction machine and the three-phase inverter. For the induction motor if the machine equations as derived in Chapter 3 are rewritten with $i_{qs}, i_{ds}\ \lambda_r$ *and* ω_r as the state variables then the Equations from 5.43 to 5.47 are obtained,

$$L_\sigma p i_{qs} = V_{qs} - r_s i_{qs} - \frac{L_m}{L_r} p\lambda_{qr} - \omega_e L_\sigma i_{ds} - \frac{L_m}{L_r}\omega_e \lambda_{dr} \tag{5.43}$$

$$L_\sigma p i_{ds} = V_{ds} - r_s i_{ds} - \frac{L_m}{L_r} p\lambda_{dr} + \omega_e L_\sigma i_{qs} - \frac{L_m}{L_r}\omega_e \lambda_{qr} \tag{5.44}$$

$$p\lambda_{qr} = \frac{-r_r}{L_r}\lambda_{qr} + \frac{r_r L_m}{L_r} i_{qs} - (\omega_e - \omega_r)\lambda_{dr} \tag{5.45}$$

$$p\lambda_{dr} = \frac{-r_r}{L_r}\lambda_{dr} + \frac{r_r L_m}{L_r} i_{ds} + (\omega_e - \omega_r)\lambda_{qr} \tag{5.46}$$

$$p\omega_r = \frac{P}{2J}\left(\frac{3P}{4}\frac{L_m}{L_r}(\lambda_{dr} i_{qs} - \lambda_{qr} i_{ds}) - T_L\right) \tag{5.47}$$

$$T_e = \frac{3P}{4}\frac{L_m}{L_r}(\lambda_{dr} i_{qs} - \lambda_{qr} i_{ds}) \tag{5.48}$$

where $L_\sigma = L_s - \frac{L_m^2}{L_r}$.

When the flux is oriented along the d-axis, $\lambda_{qr} = 0$ and $\lambda_{dr} = \lambda_r$, using this relation the above equations can be written as

$$L_\sigma p i_{qs} = V_{qs} - r_s i_{qs} - \omega_e L_\sigma i_{ds} - \frac{L_m}{L_r}\omega_e \lambda_r \tag{5.49}$$

$$L_\sigma p i_{ds} = V_{qs} - r_s i_{ds} - \frac{L_m}{L_r} p\lambda_r + \omega_e L_\sigma i_{qs} \tag{5.50}$$

$$p\lambda_r = \frac{-r_r}{L_r}\lambda_r + \frac{r_r L_m}{L_r} i_{ds} \quad (5.51)$$

$$p\omega_r = \frac{P}{2J}\left(\frac{3P}{4}\frac{L_m}{L_r}\lambda_{dr} i_{qs} - T_L\right) \quad (5.52)$$

$$p\omega_r = \frac{P}{2J}(\gamma\lambda_{dr} i_{qs} - T_L) \quad (5.53)$$

where $\gamma = \frac{3P}{4}\frac{L_m}{L_r}$.

The above Equations from 5.49 to 5.53 are used to design the controller structure using the feedback linearization technique as explained in the previous section. The slip speed is calculated and added to the rotor speed to calculate ω_e, according to the equation 5.24. In the foregoing section it is assumed that the zero sequence voltage is zero, but in the controller structure in this thesis even the zero sequence voltage cannot be assumed to be zero thus the equation for the zero sequence voltage is given as,

$$V_{os} = r_s i_{os} + L_{ls} p i_{os} \quad (5.54)$$

As can be seen Equation 5.51 and 5.53 are dependent on the d-axis and the q-axis stator current only or vice-versa. Thus controlling the rotor flux and rotor speed, a reference for the d-axis and the q-axis stator currents can be obtained. From equations 5.49 and 5.50 using the feedback linearization technique and by considering V_{qs} and V_{ds} as the outputs by controlling the currents a relation between the input and the output can be obtained.

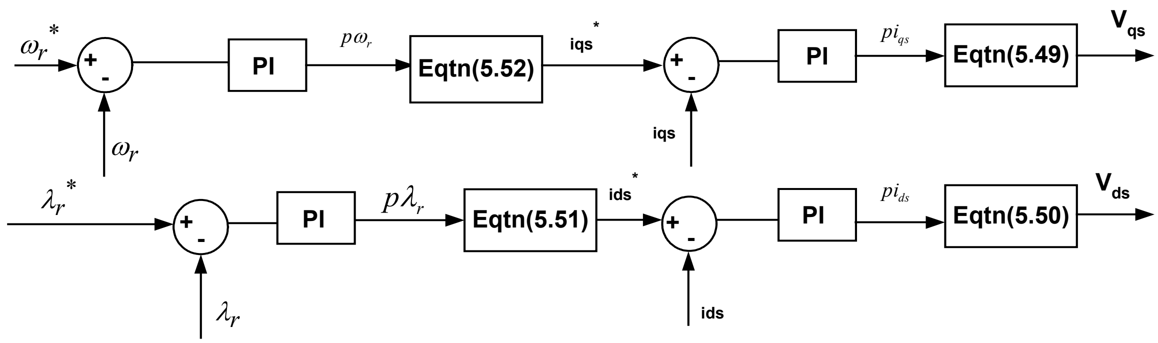

Figure 5.4: Control Diagram for the induction machine using feedback linearization technique.

Thus the control structure for the entire machine can be summarized as, controlling rotor speed and flux the reference for the stator currents is obtained, which in turn are controlled to get the reference q and d- axis voltages, these voltages when applied to the inverter synthesize the desired performance quantities. The control scheme for the induction machine part is as shown in Figure 5.4.

As can be seen from Figure 5.4, the outputs of the controller are $pi_{qs}, pi_{ds}\ p\lambda_r\ and\ p\omega_r$. The controller transfer function is that of a PI controller which is $k_p + \frac{k_i}{s}$. The input-output relation for one of the controller is derived and for the rest it is generalized.

$$(I_{qs}* - I_{qs})(k_{p1} + \frac{k_{i1}}{s}) = pI_{qs} \quad \Rightarrow \quad \frac{i_{qs}}{i_{qs}^*} = \frac{pK_{p1} + K_{i1}}{p^2 + pK_{p1} + K_{i1}} \tag{5.55}$$

Similarly,

$$\frac{i_{ds}}{i_{ds}^{*}} = \frac{pK_{p2} + K_{i2}}{p^2 + pK_{p2} + K_{i2}} \tag{5.56}$$

$$\frac{\omega_r}{\omega_r^{*}} = \frac{pK_{p3} + K_{i3}}{p^2 + pK_{p3} + K_{i3}} \tag{5.57}$$

$$\frac{\lambda_r}{\lambda_r^{*}} = \frac{pK_{p4} + K_{i4}}{p^2 + pK_{p4} + K_{i4}} \quad . \tag{5.58}$$

where $K_{p1}, K_{p2}, K_{p3}, K_{p4}$ are the proportional parts and $K_{i1}, K_{i2}, K_{i3}, K_{i4}$ are the integral parts of the PI controllers used.

After obtaining the controller transfer functions, the parameters for the PI controller are to be determined. In designing the parameters of the controller, the denominator of the transfer function is compared with Butterworth Polynomial [22]. The Butter-worth method locates the eigen values of the transfer function uniformly in the left half of the s-plane, on a circle of radius ω_o, with its center at the origin. The Butterworth polynomials for a transfer function with a second order denominator is given as:

$$p^2 + \sqrt{2} p w_0 + w_0^2 = 0 \tag{5.59}$$

Hence by comparing the denominator of the transfer function with above polynomials K_p and K_i are

$$K_p = \sqrt{2}\omega_o \qquad\qquad K_i = \omega_o^{\,2} . \tag{5.60}$$

Thus for the other controllers conditions similar to the one given above, for the PI parameters are obtained.

In this chapter a detailed analysis of the control scheme and the controller structure for the induction machine are give, with appropriate derivations. The concept of vector control for induction machine has been introduced. The feedback linearization technique for controller design has been given, and finally the parameters for the controller using the Butterworth polynomial have been discussed.