### 6.1 Introduction

Wound rotor induction machine used in variable speed WECS as shown in figures 1.4 and 3.6 is known as Double output induction generator (DOIG) because the output is fed to the grid from both the stator and rotor of the machine. The simplified diagram of the system is shown in figure 6.1.

![Diagram of Variable speed grid-connected WECS using doubly-fed wound rotor induction machine](image)

**Figure 6.1 Variable speed grid-connected WECS using doubly-fed wound rotor induction machine**

The rotor side control scheme for wound rotor induction machine is introduced in chapter 1. This arrangement offers enormous flexibility in terms of control of active and reactive powers in variable speed WECS.

To enhance the performance of variable speed WECS, control scheme for the DOIG are expected to achieve the following goals:
(1) To dynamically provide the required speed output corresponds to the maximum power capture from prescribed torque-speed curve of wind turbine.

(2) To achieve the desired power factor operation.

A vector control approach using stator flux orientation as proposed by Leonhard [67] is employed to decouple the dynamics of the active and reactive current loops in order to control the active and reactive powers. This control of active and reactive powers will lead to achieve the above goals respectively.

In this chapter a mathematical model of the doubly-fed grid-connected wound rotor induction machine and the front end converter is formulated. A design methodology is evolved for developing the current controllers. Simulation results are presented to confirm the design and modeling.

6.2 Rotor side control: Mode of operation

The operating region of the system shown in the torque speed plane is shown figure 6.2 As stated earlier, the rotor side control strategy is advantageous within a limited slip range. Hence the operating region is spread out on both sides of the synchronous speed (Ws) implying both sub and super synchronous modes of operation. Moreover the machine can operate in the motoring and generating modes irrespective of the speed. Thus four distinct modes of operation can be achieved though rotor side control corresponding to the four quadrants in the torque speed plane.

In figure 6.2, mode 1 refers to positive torque and sub synchronous speed and is termed as sub synchronous motoring (i.e normal motoring) operation.
Mode 2 corresponds to positive torque and super synchronous speed and is called super synchronous motoring. Similarly, mode 3 corresponds to sub synchronous generating and mode 4 to super synchronous generation. The following section describe how these different modes of operation can be achieved through rotor side control.

![Operating region of the doubly fed induction machine with rotor side control](image)

**Figure 6.2 Operating region of the doubly fed induction machine with rotor side control**

![Approximate equivalent circuit with rotor current control](image)

**Figure 6.3 Approximate equivalent circuit with rotor current control**

**Mode 1: Sub synchronous motoring**

A simplified equivalent circuit of the doubly fed wound rotor induction machine controlled from the rotor side is shown in figure 6.2. It is assumed that the rotor currents can be injected at any desired phase, frequency and magnitude. Therefore, the rotor circuit can be represented by a controllable
current source. The equivalent circuit is drawn in the stator reference frame hence the rotor current is represented as $i_r$. The steady-state phasor diagram and power flow for the sub synchronous motoring mode of operation are shown in figure 6.4.

Neglecting the stator resistance, it may be assumed that the stator flux ($\psi_s$) remains constant in magnitude and frequency since the stator is connected to the power grid. $\psi_s$ has two components; the stator leakage component and the magnetizing component. The former is due to the stator current alone, while the latter is due to both the stator and rotor currents. An equivalent current ($I_{ms}$) can be defined in the stator reference frame, which is responsible for the stator flux. This is termed as the stator flux magnetizing current [67]. The direction of $\psi_s$ (which is in phase with $I_{ms}$) is defined as the d-axis and the direction of the stator voltage, which is at quadrature of $\psi_s$, is termed as the q-axis. It is possible to resolve $i_s$ and $i_r$ along and perpendicular to $I_{ms}$ (The components of the currents along the d-axis are represented with subscript d, and those along the q-axis with subscript q).
Figure 6.4 Phasor diagram and power flow diagram during subsynchronous motoring
Since $\psi_s$ is constant, it implies the $i_{ms}$ is also constant and equals sum of $i_{sd}$ and $i_{rd}$. With current control being exercised in the rotor circuit, an injection of positive $i_{rd}$ will naturally result in a lesser value of $i_{sd}$ being drawn from the stator terminal. The stator power factor is thereby improved. This feature is clearly depicted in figures 6.4(b). Figure 6.4 (a) shows $i_{ms}$ being fully supplied from the stator side, as in the case of cage rotor induction machine, whereas in figure 6.4(b) it is partially supplied from the rotor side and partly from the stator side. It may be noted here that $i_{rd}$ will never be made negative. This would mean that the stator has to supply the magnetizing energy of the machine as well as the reactive energy demand of the rotor circuit bringing down the stator power factor to a very low value.

Along the q-axis, the magnitude of the active component of stator current($i_{sq}$) is directly proportional to $i_{rq}$, but opposite in sign. In fact, the induction machine can be looked upon as current transformer as far as the active power flow in the stator and rotor circuit is concerned. Hence, to produce a motoring torque (i.e. positive torque), $i_{rd}$, has to be negative. This is evident from figure 6.4, a negative $i_{rd}$ induces a positive $i_{sq}$, implying flow of active power into the stator circuit. Below the synchronous speed the rotor falls behind the air-gap flux and the rotor introduced emf ($e_r$) the lags the mutual flux ($\psi_m$) by $90^0$ as shown in figures 6.4 (a) and (b). The locus of $i_s$ and $i_r$ for constant active power flow is shown in figure 6.4c. As the tip of the for the rotor current phasor is moved from B to A, the stator current phasor moves in the opposite direction from B’ to A’. From this phasor diagram it may be appreciated that some amount power
supplied from the rotor side is more than the machine requirement. This is, however possible when the active load demand is low and there is adequate current margin in the rotor coils. In order to utilize the copper in the stator and rotor circuits effectively, it is advisable to divide the reactive power demand between the two ports.

Under the condition of subsynchronous motoring the stator voltage phasor \( U_s \), leads the air-gap voltage \( e_s - e_r \) under all conditions of load which indicates power flowing into the stator. Also the rotor current \( i_r \) makes an angle less than \( 90^\circ \) with \( e_r \), the rotor-included emf, implying that active power is being extracted from the rotor circuit. This rotor power, or the slip power, is recovered from the rotor circuit and fed back to the mains, thereby increasing the system efficiency. The mechanical power output is roughly the difference between the stator and rotor powers (figure 6.4 (d)).

**Mode 2 : Supersynchronous motoring**

With \( i_{rq} \) remaining negative if the machine runs above synchronous speed, it enters the supersynchronous motoring mode of operation. The rotor now moves ahead of the air-gap flux \( \psi_m \) and, therefore, \( e_r \) leads \( \psi_m \) by \( 90^\circ \). The phase relations between the stator and rotor currents remain as in mode 1; only the direction of rotor power reverses as \( i_r \) now makes an angle more than \( 90^\circ \) with \( e_r \) (figure 6.5).
It may be noted in this mode of operation, if the stator input power is 1 p.u. and the motor is at a slip of s p.u., the mechanical output that can be obtained is (1+s) p.u. which is more than the rating of the machine.

Mode 3: Sub synchronous generation

If a positive \( i_{r_q} \) is injected into the rotor circuit, \( i_{sq} \) changes direction and becomes negative. Therefore, the active power flow into the stator becomes negative indicating that the machine is generating. This can also be appreciated from the fact that the stator terminal voltage vector \( (u_s) \) now lags the stator-included emf. The phase angle between \( i_r \) and \( e_r \) exceeds 90°, implying that power is fed into the rotor circuit. The power flow and phasor diagrams are given in figure 6.6.
Mode 4: super synchronous generation

With $i_{rq}$ remaining positive, the machine can go over to the super synchronous generating mode. As far as the stator circuit is concerned everything remains the same as in mode 3; only the rotor power flow changes its direction. With the rotor-included emf $e_r$ leading the air-gap flux, the angle between $i_r$ and $e_r$ becomes less than $90^0$ indicating power flow out of the rotor. It is interesting to note that in super synchronous generation mode the shaft power is recovered from both the stator and the rotor ends. Therefore, if 1 p.u. power is extracted from the stator, while the machine is running at a slip $s$, the total power generated will be $(1+s)$ p.u. Hence in the super synchronous generation mode it
is actually possible to generate power that is more than the rating of the machine (figure 6.7).

![Phasor diagram and power flow diagram during super synchronous generation](image)

**Fig 6.7 Phasor diagram and power flow diagram during super synchronous generation**

### 6.3 Machine Model in Field Coordinates

In a doubly-fed wound rotor induction machine, control is exerted on the rotor side while the stator remains connected to a constant voltage constant frequency source. In order to formulate the dynamic modeling in the field coordinates, it is assumed that the rotor side converter is equipped with fast-acting current loops. Hence, given a reference $i_{r}^{*}$, the rotor current space phasor $i_{r}$ follows it within a finite but extremely short interval of time. The rotor voltage equation is used for designing the rotor current controller, as discussed in a later section. For the present, the rotor side can be simply represented by a
controllable current source [figure 6.8] and the rotor current phasor can be taken as an input to the machine model. It is the stator voltage equation, which determines the dynamic behavior of the machine. This equation, in the stationary reference frame, is furnished below.

\[
R_s i_s + (1 + \sigma_s) L_o \frac{d}{dt} i_s + L_o \frac{d}{dt} (i_s e^{j\theta}) = u_s
\]

\[
R_s i_s + L_o \frac{d}{dt} [(1 + \sigma_s)i_s + i_r e^{j\theta}] = u_s
\]

(6.1)

![Figure 6.8 Equivalent circuit in stator reference frame](image)

The magnetizing current vector is defined as

\[
i_{ms} = (1 + \sigma_s) i_s + i_r e^{j\theta}
\]

\[
= (1 + \sigma_s)i_s + s_i r
\]

(6.2)

\(i_{ms}\) is the equivalent current vector in the stator reference frame responsible producing the stator flux \(\psi_s\) as depicted in the equivalent circuit of figure 6.8. Hence it may be called the stator flux magnetizing current. In terms of \(I_{ms}\) equn.6.1 can be written as
\[ R_s i_s + L_o \frac{d}{dt} i_{ms} = u_s \]  

(6.3)

Since the quantity over which direct control can be exercised is \( i_s \) (and not \( i_r \)), Equn.6.3 needs to be expressed in terms of the rotor currents.

Substituting for \( i_s \) in Equn.6.3, using Equn.6.2 we get

\[ R_s \left[ \frac{i_{ms} - i_r e^{j\omega}}{1 + \sigma s} \right] + L_o \frac{d}{dt} i_{ms} = u_s \]

or,

\[ T_s \frac{d}{dt} i_{ms} + \frac{i_{ms} - i_r e^{j\omega}}{1 + \sigma s} = \frac{(1 + \sigma s)}{R_s} u_s + i_r e^{j\omega} \]  

(6.4)

where \( T_s = \frac{L_o (1 + \sigma s)}{R_s} = \frac{L_s}{R_s} \) is the electrical time-constant of the stator circuit.

The above equation Equn.(6.4) is defined in the stator reference frame. It can now be expressed in terms of a coordinate system fixed to the stator flux \( \psi_s \) or equivalently to the current \( i_{ms} \). In order to do this \( i_{ms} \) is first expressed in polar from with respect to stator coordinates as

\[ i_{ms} = i_{ms} e^{j\mu} \]  

(6.5)

where \( i_{ms} \) is the instantaneous magnitude of the current space phasor \( i_{ms} \) and \( \mu \) is its instantaneous position with respect to the stationary axis. The various phase relationship are shown in figure 6.9.
Now Eqn.(6.4) can be written as

\[ T_s \frac{d}{dt} (i_{ms} e^{j\mu}) + i_{ms} e^{j\mu} = \frac{1+\sigma_s}{R_s} u_s + i_r e^{j\mu} \]

or, \( T_s \frac{di_{ms}}{dt} e^{j\mu} + T_s i_{ms} (j \frac{d\mu}{dt}) e^{j\mu} + i_{ms} e^{j\mu} = \frac{1+\sigma_s}{R_s} u_s + i_r e^{j\mu} \)  \hspace{1cm} (6.6)

In the field coordinates, the stator voltage and rotor current space phasors can be represented as

\[ u_s e^{-j\mu} = u_{sd} + ju_{sq} \]

and \( i_r e^{j(\omega_e)} = i_{rd} + ji_{rq} \)  \hspace{1cm} (6.8)

substituting this in equn.6.7 and separating the real and imaginary parts yields the following equations.

\[ T_s \frac{di_{ms}}{dt} + jT_s i_{ms} \frac{d\mu}{dt} + i_{ms} = \frac{1+\sigma_s}{R_s} u_s e^{-j\mu} + i_r e^{j(\omega_e)} \]  \hspace{1cm} (6.7)
\[
T_s \frac{d i_{ms}}{dt} + i_{ns} = \frac{1 + \sigma_s}{R_s} u_{sd} + i_{rd}
\]  \hspace{1cm} (6.9)

\[
\frac{d \mu}{dt} = \omega_{ms} = \frac{1}{T_s i_{ms}} \left[ \frac{1 + \sigma_s}{R_s} u_{sq} + i_{rq} \right]
\]  \hspace{1cm} (6.10)

Eqns.6.9 and 6.10 represent the dynamics of the field vector magnitude and angle respectively. The stator voltage and rotor current vectors are the two inputs to the system, of which the former is not controllable (hence a disturbance variable) and the latter is the control variable.

The electromagnetic torque developed can also be expressed in terms of the current vector. In the stationary coordinates the torque equation is given by

\[
m_d = \frac{2}{3} p \frac{L_o}{2} \Im \left\{ i_{ms} (i e^{j \theta})^* \right\}
\]  \hspace{1cm} (6.11)

Substituting for \(i_s\) using Equn.6.2, we get

\[
m_d = \frac{2}{3} p \frac{L_o}{2} \Im \left\{ \frac{(i_{ms} - s i_{ms})}{1 + \sigma_s} i_{rd}^* \right\}
\]  \hspace{1cm} (6.12)

The complete set of equations that describe the machine dynamics in the field coordinates can be therefore, written as
\[
T_s \frac{di_{ms}}{dt} + i_{ms} = \frac{1 + \sigma_s}{R_s} u_{sd} + i_{nd}
\] (6.13 a)

\[
\frac{d\mu}{dt} = \omega ms = \frac{1}{T_s i_{ms}} \left[ \frac{1 + \sigma_s}{R_s} u_{sq} + i_{rq} \right]
\] (6.13 b)

\[
J \frac{d\omega}{dt} = -\frac{2 P}{3} \frac{L_o}{2} \frac{1 + \sigma_s}{1 + \sigma_s} i_{ms} i_{rq} - m_1
\] (6.13 c)

\[
\frac{de}{dt} = \frac{p}{2} \omega = \omega_c
\] (6.13 d)

The simulation block diagram of the doubly-fed wound rotor induction machine modeled in the field coordinates is given in figure 6.10. The inputs to the system are the stator voltage and the rotor currents. It is assumed that there is a controlled current source in the rotor circuit which is capable of injecting currents at appropriate phase, frequency and magnitude. The rotor currents are first transformed to the stator reference frame using the operator \( e^{j\mu} \). Subsequently the stator voltages and rotor currents (in the stator reference frame) are transformed to the field coordinates by multiplying with \( e^{-j\mu} \). The angle \( \mu \) is derived by solving the q-axis equation equn. (6.13b). The magnitude of \( i_{ms} \) is computed using the d-axis equation equn. (6.13a). It may be noted that the alternating quantities in the stationary coordinates, when transformed to the field coordinates appear as dc quantities.
6.4 Field Oriented Control

The rotor circuit consists of a three phase voltage source inverter operating in the current-controlled mode. The stator is connected to a constant magnitude, constant frequency source which has ideally infinite capacity for sourcing and sinking active and reactive powers. It may also be assumed for the time being that the dc source for the inverter can supply or sink the active and reactive powers handled by the rotor without affecting the dc bus voltage. (In practice the FEC interfaces the dc bus with the ac grid, which is discussed in the later section). The schematic block diagram of this arrangement is shown in figure 6.11.
6.4.1 Rotor Equation in Field Coordinates

For designing the rotor current controller the rotor voltage equation has to be considered. This equation is

\[
R_r i_r + L_r \frac{d}{dt} i_r + L_o \frac{d}{dt} (i_s e^{-j \omega t}) = u_r
\]  

(6.14)

is in the rotor reference frame and needs to be transformed to the field coordinates for field-oriented control. Moreover, the equation has to be
expressed only in terms of $i_r$, the variable to be controlled, and $i_{ms}$, the field variable.

From eqn. (6.2)

$$i_s = \frac{i_{ms} - i_r e^{j \xi}}{1 + \sigma_s} \quad (6.15)$$

Substituting for $i_s$ using eqn.(6.15), eqn.(6.14) may be expressed as

$$R_r i_r + L_r \frac{d}{dt} i_r + L_o \frac{d}{dt} \left[ \frac{i_{ms} e^{-j \xi} - i_r}{1 + \sigma_s} \right] = u_r \quad (6.16)$$

Now,

$$\frac{L_o}{1 + \sigma_s} = \frac{L_o^2 L_r}{L_r L_r} = (1 - \sigma) L_r$$

and $i_{ms} = i_{ms} e^{j \mu}$

Therefore, eqn. (6.16) can be written as

$$R_r (i_{rd} + j i_{rq}) e^{j (\mu - \xi)} + \sigma L_r \frac{d}{dt} [(i_{rd} + j i_{rq}) e^{j (\mu - \xi)}]$$

$$+ (1 - \sigma) L_r \frac{d}{dt} [i_{ms} e^{j (\mu - \xi)}]$$

$$= (u_{rd} + j u_{rq}) e^{j (\mu - \xi)}$$

or,

$$R_r (i_{rd} + j i_{rq}) e^{j (\mu - \xi)} + \sigma L_r \frac{d}{dt} [(i_{rd} + j i_{rq}) e^{j (\mu - \xi)}]$$

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\[ + \sigma L_r \frac{d(\mu - \varepsilon)}{dt}(i_{nl} + j i_{nq})e^{j(\mu - \varepsilon)} + (1 - \sigma)L_r \frac{di_{ms}}{dt}e^{j(\mu - \varepsilon)} \]

\[ + (1 - \sigma)L_r i_{ms} j \frac{d(\mu - \varepsilon)}{dt}e^{j(\mu - \varepsilon)} \]

\[ = (u_{nd} + ju_{nq})e^{j(\mu - \varepsilon)} \quad (6.18) \]

In order to transform this equation to the field-oriented reference frame, both sides have to be multiplied by \( e^{-j(\mu - \varepsilon)} \). Thereby the following complex equation can be derived.

\[ R_r (i_{nl} + j i_{nq}) + \sigma L_r \frac{d}{dt}(i_{nl} + j i_{nq}) + j(\omega_{ms} - \omega_c)\sigma L_r (i_{nl} + j i_{nq}) \]

\[ + (1 - \sigma)L_r \frac{di_{ms}}{dt} + j(\omega_{ms} - \omega_c)(1 - \sigma)L_r i_{ms} \]

\[ = (u_{nd} + ju_{nq}) \quad (6.19) \]

where, \( \frac{d(\mu - \varepsilon)}{dt} = \omega_{ms} - \omega_c \) is the slip frequency \( (6.20) \)

Finally, the real and the imaginary parts are separated to get the d-axis and q-axis equations respectively as given below.

\[ \sigma Tr \frac{di_{nl}}{dt} + i_{nl} = \frac{\mu_n}{R_r} + (\omega_{ms} - \omega_c)\sigma T_r i_{nq} - (1 - \sigma)T_r \frac{di_{ms}}{dt} \quad (6.21) \]

\[ \sigma Tr \frac{di_{nq}}{dt} + i_{nq} = \frac{\mu_n}{R_r} - (\omega_{ms} - \omega_c)\sigma T_r i_{nl} - (\omega_{ms} - \omega_c)(1 - \sigma)T_r i_{ms} \quad (6.22) \]

These equations represent the dynamics of the rotor currents in the field coordinate system. It is observed that due to the presence of the rotational emf terms, there is some amount of cross-coupling between the d and q axes.
However, the current-loop dynamics in the two axes can be made independent of each other by compensating for these cross-coupling terms.

It is interesting to note the connotations of the terms underlined in these two equations. Multiplying these terms with $R_r$ gives the corresponding voltages, which can be interpreted as follows.

(a) $(\omega_{ms} - \omega_e)\sigma T_r i_{m}\times R_r = (\omega_{ms} - \omega_e)\sigma L_r i_{m}$: This is the rotational emf induced in the d axis due to the q axis rotor current. Since the relative speed of the rotor with respect to the field axis is $(\omega_{ms} - \omega_e)$, the frequency term involved in this equation corresponds to the slip frequency.

(b) $-(1 - \sigma)T_r \frac{d}{dt} i_m x R_r = -(1 - \sigma) L_r \frac{d}{dt} i_m$: This denotes the transformer induced voltage in the d-axis due to the field current. Obviously this term will not appear in the q-axis.

(c) $-(\omega_{ms} - \omega_e)\sigma T_r i_{rd}\times R_r = -(\omega_{ms} - \omega_e)\sigma L_r i_{rd}$: This term gives the rotationally induced emf in the q-axis due to the d-axis current, similar to the first term.

(d) $(\omega_{ms} - \omega)(1 - \sigma)T_r i_{me}\times R_r = (\omega_{ms} - \omega_e)(1 - \sigma)L_r i_{me}$: This denotes the speed emf induced in the q-axis due to the field. The frequency term involved in this equation is again the slip frequency.

Due to the presence of these terms, there exists some coupling between the two axes. However, as the slip range is limited, the contribution of the terms (a) and (c) is relatively small compared to the speed emf term (d). The transformer emf term (b) also does not exist after the flux has built up, provided the stator voltage is constant in magnitude and frequency. While designing the
rotor current controller it is possible to compensate for these terms and make the loop dynamics in the two axes independent of each other. This is discussed in the following section.

6.4.2 Design of Rotor Current Controller in Field Coordinates

It is obvious that if the rotor current needs to be controlled in the field coordinates, two independent controllers are needed; one for the d-axis and the other for q-axis. The design method is same for both; only the feed forward terms differ in each case. Design of a proportional (p) controller, and a proportional-integral (PI) controller are presented here.

(a) Proportional Controller

Let the desired current loop dynamics in the d-axis be given by

\[
T_{ir} \frac{di_{rd}}{dt} + i_{rd} = \frac{i^*_{rd}}{K_{ir}} \tag{6.23}
\]

Where \( T_{ir} \) is the desired current-loop time constant and \( K_{ir} \) is the current sensor gain.

The task is now to find out the d-axis component of the instantaneous inverter terminal voltage required to produce the current dynamics given by eqn.(6.23).

Substituting for \( \frac{di_{rd}}{dt} \) from eqn.(6.23) in eqn.(6.21) gives

\[
u_{rd} = \frac{\sigma L_r}{T_{ir} K_{ir}} (i^*_{rd} - K_{ir} i_{rd}) + R_r i_{rd} + \left(1 - \sigma\right) L_r \frac{di_{ms}}{dt} - \left(\omega_m - \omega_c\right) \sigma L_r i_{rq} \tag{6.24}
\]
The inverter can be modeled as a gain block $G_r$. For sine-triangle modulation the inverter gain depends on the dc bus voltage $u_{dc}$ and the peak of the triangle $u_{tri}$.

$$G_r = \frac{u_{dc}}{2u_{tri}} \quad (6.25)$$

In order to make the inverter gain constant, the peak of the carrier triangular waveform $u_{tri}$ is made proportional to $u_{dc}$.

Therefore the reference for the d-axis component of rotor voltage is given by

$$u^{*}_{rd} = \frac{u_{rd}}{G_r}$$

$$\frac{\sigma L_r}{T_r K_r G_r} (i^{*}_{rd} - K_{ir} i_{rd}) + \frac{R_r}{G_r} i_{rd} + \frac{(1-\sigma)L_r}{G_r} \frac{di_{ms}}{dt} - \frac{(\omega_{ms} - \omega_c)\sigma L_r}{G_r} i_{rq} \quad (6.26)$$

Assuming same current loop dynamics for the q-axis the reference for the q-axis component of the rotor voltage can be expressed as

$$u^{*}_{rq} = \frac{u_{rq}}{G_r}$$

$$= \frac{\sigma L_r}{T_r K_r G_r} (i^{*}_{rq} - K_{ir} i_{rq}) + \frac{R_r}{G_r} i_{rq}$$

$$+ \frac{(\omega_{ms} - \omega_c)(1-\sigma)L_r}{G_r} i_{ms} + \frac{(\omega_{ms} - \omega_c)\sigma L_r}{G_r} i_{rd} \quad (6.27)$$

If the impressed rotor voltages are in accordance with eqn. (6.26) and eqn.(6.27), the field and quadrature axes rotor currents can be controlled independently. The rotor current controller, therefore, does not contribute significantly to the dynamics of the system. It only ensures that the rotor currents
track the reference signals produced by the outer loops. It may be appreciated that the current loop dynamics can be made much faster than the rotor time constant. However, a practical limitation to the bandwidth of the current controller is imposed by the switching frequency. Since the controller is designed in the field coordinates, all the quantities are dc and implementation of the controller becomes simpler.

The simulation block diagram of the rotor current controller is shown in four parts. Figure 6.12 shows the computation of the vector and transformation of the rotor currents to the field coordinates. Assuming that at \( t = 0 \), the rotor and stator axes are aligned, the rotor position \( \varepsilon \) can be directly obtained in the simulation by integrating the shaft speed. The rotor currents are first transformed to the stationary coordinates with this angle information \( \varepsilon \). The angle \( \mu \), which the field axis makes with the stator coordinates, as well as the magnitude of \( i_{ms} \), can be calculated from the stator and rotor currents as follows.

In the stator coordinates,

\[
i_{ms} e^{j\mu} = (1 + \sigma_s)(i_{sa} + j i_{sb}) + (i_{ra} + j i_{rb})(\cos \varepsilon + j \sin \varepsilon)
\]

\[
= [(1 + \sigma_s)i_{sa} + i_{ra} \cos \varepsilon - i_{rb} \sin \varepsilon]
\]

\[
+ j[(1 + \sigma_s)i_{sb} + i_{ra} \sin \varepsilon + i_{rb} \cos \varepsilon]
\]

Therefore \( i_{ms} = \left[\left(1 + \sigma_s\right)i_{sa} + i_{ra} \cos \varepsilon - i_{rb} \sin \varepsilon\right]^2 \) \( + \left[\left(1 + \sigma_s\right)i_{sb} + i_{ra} \sin \varepsilon + i_{rb} \cos \varepsilon\right]^2 \right]^{1/2} \) \( \mu = \arctan\left[\frac{(1 + \sigma_s)i_{sb} + i_{ra} \sin \varepsilon + i_{rb} \cos \varepsilon}{(1 + \sigma_s)i_{sa} + i_{ra} \cos \varepsilon - i_{rb} \sin \varepsilon}\right] \) \( (6.27) \)

\( (6.28) \)

\( (6.29) \)
The d-axis controller block diagram along with the plant is given in figure 6.13(a). All the parameters necessary for constructing the feed forward terms have already been computed in the previous stage. After generation of the reference voltages $u_{rd}^*$ and $u_{rq}^*$, they are again transformed back to the rotor reference frame by multiplying with the inverse transformation operator $e^{j(\mu-\epsilon)}$. However, for the sake of simplification, transformation of $u_{rd}^*$ from the field coordinates to the rotor reference frame in the controller, and the corresponding forward transformation in the machine model from this diagram. It is evident from the diagram that the controller just adds or subtracts the disturbance inputs to the machine model reducing the plant to merely an integrator. Hence, the closed loop system behaves like a first-order lag circuit whose time-constant can be modified by changing the proportional gain of the system, as figures 6.13(b) and 6.13(c). The q-axis current controller can be similarly modeled; the compensating
terms will only be different in this case. In the controller block diagram the current sensor gain $K_{ir}$ is assumed to be unity for simplifying the diagrams.

Figure 6.13 Block diagram of d-axis proportional rotor current controller

(b) Proportional Integral Controller

In the proportional controller, the steady-state error in the rotor currents will depend on the accuracy of computation of the feedforward terms. Any error in the compensation terms would result in slight modification of the dynamic response. In the practical implementation, it is extremely difficult to perfectly nullify the disturbance terms owing to measurements and computational errors. Therefore a proportional integral controller needs to be incorporated. This is particularly important when the machine is run in the torque-control mode without any outer speed loop. The flux computation and transformations shown in
figure 6.12 remain identical. The plant is modeled as a first-order lag system with the two rotational emfs as disturbance inputs as shown in figure 6.14(a). These two terms are canceled through feed forward compensation as before. In the controller, the PI-time-constant is made equal to $\sigma T_r$, so that the dynamics of the system is decided by the proportional gain, as shown in figure 6.14(b) and figure 6.14(C).

The choice of proportional gain follows from the equation $K_{pir} = \frac{\sigma T_r}{T_{ir}} R_r$, where $T_{ir}$ is the desired effective time constant of the current loop.

![Figure 6.14 Block diagram of d-axis proportional-integral rotor current controller](image-url)
6.5 Simulation Results-Rotor Side Control

The entire system is simulated on the MATLAB (6.0)-SIMULINK Platform. The simulation model comprises different functional modules or subsystems. Each of these modules, in turn have several levels of subsystems which are developed using the standard SIMULINK library.

Figure 6.15 SIMULINK model of the doubly-fed SRIM

As an example, the modeling of the doubly-fed SRIM may be considered. The machine model consists of the transformation blocks, electrical subsystem, torque converter block and, the mechanical subsystem. This is shown in figure 6.15. The transformation blocks include 3 phase to 2 phase transformation, rotor coordinate to stator coordinate transformation and, the corresponding inverse transformations (used in the current inspection block.). The electrical subsystem is modeled in the stationary coordinate system. The state space equations
(equn. (A32)) to compute the stator and rotor currents in the stator reference frame are derived in Appendix A. The torque equation, given by equn. (A33), is executed in the torque converter block. Finally, the mechanical subsystem computes the machine speed from the mechanical dynamics, as given by equn. (A22d). The position of the rotor with respect to the stator is obtained through integration of the shaft speed in the \( \sin \psi \), \( \cos \psi \) block. These modules with appropriate interconnections are grouped together to form the SRIM block in the system simulation model of figure 6.16. In a similar manner, the other functional modules of figure 6.16 are developed.

Figure 6.16 SIMULINK block diagram of a speed-controlled drive using doubly-fed SRIM
A speed controlled drive using grid-connected doubly-fed SRIM is simulated. Stator flux orientation, as discussed in the earlier sections, is employed. The d-axis and q-axis current controllers are designed in the field reference frame. For the speed loop, a PI controller is employed which generates the q-axis / active current reference $i^*_{rq}$. The d-axis reference $i^*_{rd}$ set in open loop.

The speed response of the drive under no load is given in figure 6.17 (a). The motor is assumed to be started DOL with the rotor shorted. At $t=0.25s$, the rotor side control is released with a speed reference $\omega^* = 0.75 \text{ p.u.}$ At $t=1.25s$, the speed reference is given a step change from 0.75 p.u. to 1.25 p.u. The corresponding motor torque and $i_{rq}$ are given in figure 6.17(b) and figure 6.17(c) respectively. The speed controller time constant is set to 100 ms. At $t=1.75s$, $i^*_{rd}$, is given a step change to 0.75 p.u. This results in transfer of the reactive power from the stator to the rotor side. However, a change in $i_{rd}$ does not affect $i_{rq}$, as can be seen from these plots.

The dynamic response of the current loops are given in figure 6.18 and figure 6.19. The q-axis current loop time constant is designed for 1ms and, the d-axis loop time constant is designed for 4ms. (The d-axis reference is not required to be varied dynamically; so the dynamic response of $i_{rd}$ need not be very fast). The actual stator currents and, the rotor currents in the stator reference frame are also shown. In figure 6.18 $i_{rd}$ is zero, so the stator supplies the reactive power and the rotor power factor is unity. In figure 6.19, the reactive power is transferred from the stator to the rotor side, resulting in an improvement in the stator power factor.
Figure 6.17(a) Simulated speed response of the speed-controlled grid-connected SRIM drive

Figure 6.17(b) Simulated torque response of the speed-controlled grid-connected SRIM drive
Figure 6.17(c) Simulated $i_q$ response of the speed-controlled grid-connected SRIM drive

Figure 6.17(d) Simulated $i_d$ response of the speed-controlled grid-connected SRIM drive
Figure 6.18(a) Simulated step response of $i_{rq}$ for the grid-connected SRIM

Figure 6.18(b) Corresponding simulated response of $i_{rd}$ for the grid-connected SRIM
Figure 6.18(c) Corresponding simulated response of $i_s$ along with $u_s$ for the grid-connected SRIM

Figure 6.18(d) Corresponding simulated response of $i_r$ along with $u_s$ for the grid-connected SRIM
Figure 6.19(a) Simulated step response of $i_{rd}$ for the grid-connected SRIM

Figure 6.19(b) Corresponding simulated response of $i_{rq}$ for the grid-connected SRIM
Figure 6.19(c) Corresponding simulated response of $i_s$ along with $u_s$ for the grid-connected SRIM

Figure 6.19(d) Corresponding simulated response of $i_r$ along with $u_s$ for the grid-connected SRIM
6.6 Front end Converter

A conventional phase-controlled rectifier has several serious disadvantages. Firstly, with the dc bus polarity remaining constant, power flow can be in one direction only. Secondly, it draws reactive power from the line, which is substantial at large firing angles. Lastly, the current drawn from the mains is far from sinusoidal. Obviously, a four-quadrant drive or generating system cannot be implemented with a phase-controlled converter in the line side.

The front end converter employs a three-phase inverter bridge topology and is controlled to enable power flow in both directions, keeping the dc bus voltage within good regulation. It can be operated at any desired power factor, and hence, can even act as a reactive power sources as far as the grid is concerned. The converter is operated as a PWM voltage source inverter in the current-controlled mode; so, the harmonics in the line current waveform are substantially reduced.

It is understood that by employing stator voltage orientation, the active and reactive current at the input of the front end converter can be controlled in the synchronous reference frame. The control essentially has the same structure as the rotor side controller. The objectives of control and, modeling of the power and control circuit are described in the following sections.
6.7 System Description

The schematic block diagram of the front end converter is shown in figure 6.20. The transformer in the input side is used to match the voltage levels between the dc bus and the ac side. The rotor side converter operates within a limited frequency range. Hence, the dc bus voltage requirement is less when compared to the stator side control schemes of cage rotor induction machine. The saving in the converter rating in the rotor side is achieved due to reduction in voltage rating of the power devices. With a reduced dc bus voltage it, however, becomes necessary to use a transformer at the input of the front end converter. Since the rotor side need not be isolated from the ac grid, an autotransformer can be used instead. This reduces the cost and weight of the equipment considerably.

The PWM switching converter is connected to the secondary side of the transformer through series chokes. These inductors act as buffers between the
two voltage sources. The choice of the values for these inductors depends on the switching frequency, allowable harmonics in the input current wave and the reactive power requirement.

The objectives for the control of the converter are,

i) Voltage regulation of the dc bus 

ii) bi-directional power flow, 

iii) operation at any desired power factor, and 

iv) low current harmonics.

6.8 Principle of Operation and Control

The front end converter requires closed-loop control to meet the stated objectives. The basic strategy for control and resulting circuit behavior can be explained easily by means of the phasor diagrams given in figure 6.21.

![Diagram](image)

**Figure 6.21 Equivalent circuit and phasor diagrams for the front end converter**
The primary objective of control is dc bus voltage regulation. A change in the dc bus voltage can be attributed to an imbalance between the active powers between the ac and dc sides. (The effect of reactive power on the dc bus is to produce ripples in the voltage even though the average value remains the same.) Hence, the voltage error in the dc side is an indication of the active power demand in the ac side. If the demand is positive, active power drawn from grid needs to be increased; if the demand is negative, it has to be fed back to the grid. Since the converter has bidirectional switches, current flow can be in either direction and it is possible to source or sink active power in the ac side.

Figure 6.21(a) shows the single phase equivalent of the ac side of the front end converter. If the load current in the dc side for a given dc bus voltage is known, the current drawn from the ac side at any desired power factor can be calculated applying power balance between the ac and dc sides. Consequently, subtracting the reactive drop from the source voltage, the magnitude and phase of the inverter terminal voltage with respect to the source can be computed. The inverter, therefore, acts as a fixed frequency source with controllable phase and magnitude. Figures 6.21(b) and 6.21(d) represent steady-state phasor diagrams at unity power factor operation when power is flowing from ac to dc side and vice versa. The corresponding phasor diagrams for leading power factor operation are illustrated in figures 6.21(c) and 6.21(e). It is observed that the magnitude of the inverter terminal voltage increases in this case. The amount of reactive power that can be injected into the grid depends on the available dc bus voltage and the value of per unit inductance in the ac side.
The terminal voltage of the inverter will also contain switching harmonics apart from the fundamentals. As far as the harmonics are concerned the ac source acts as a short-circuit and the effective impedance to the harmonic current is $nX_{fe}$. For high frequency switching (more than 1kHz), the harmonic impedance is quite high resulting in very low distortion of the ac side current waveform.

With the above scheme of control it is possible to achieve the desired objectives as stated earlier. The salient features of the control strategy can be summed up as the following.

- Employs an outer voltage control loop to regulate the dc bus voltage and an inner current control loop to control the ac side inductor current.
- Outer voltage loop decides the value of inductor current to meet the active power balance between the two dc and ac sides.
- The current loop tracks this reference by adjusting the inverter terminal voltage so that proper phase relationship between the supply voltage and the inductor current is maintained for a given power factor operation.
- Employs current-controlled sinusoidal PWM for current tracking.
- The current controller is designed in synchronously rotating reference frame with orientation being done with respect to the supply voltage space phasor.

Figure 6.22 shows the schematic block diagram of the control structure of the front end converter. The voltage controller is a proportional-integral controller with feed-forward of the load current and generates $i_{feq}$ or the references for the
active component of current. The reference for the reactive component of current $i_{\text{fed}}$ is set in open loop; e.g. it is set to zero for unity power factor operation. The current controller operates in the synchronous reference frame and generates $u_{\text{fed}}^*$ and $u_{\text{feq}}^*$ for the inverter terminal voltage. These references are first transformed back to the stationary reference frame, and then from two phase to three phase quantities. Finally, the three phase references are compared with a triangular carrier to generate the PWM signals for the inverter switches. The modeling of the power circuit and design of the controllers are discussed in detail in the following sections.

Figure 6.22 Schematic block diagram of the control structure of the front-end converter
6.9 Modeling of the Power Circuit

In the stationary reference frame, the ac side voltage equations for the three phases can be written as follows.

\[ u_{ac1} = u_{fe1} + L_{fe} \frac{d}{dt} i_{fe1} \]  \hspace{1cm} (6.30)

\[ u_{ac2} = u_{fe2} + L_{fe} \frac{d}{dt} i_{fe2} \]  \hspace{1cm} (6.31)

\[ u_{ac3} = u_{fe3} + L_{fe} \frac{d}{dt} i_{fe3} \]  \hspace{1cm} (6.32)

The above equations can also be represented in terms of space phasors as

\[ u_{ac} = u_{fe} + L_{fe} \frac{d}{dt} i_{fe} \]  \hspace{1cm} (6.33)

where, \[ u_{ac} = u_{fe} + L_{fe} \frac{d}{dt} i_{fe} \]  \hspace{1cm} (6.34)

\[ u_{ac\beta} = u_{fe\beta} + L_{fe} \frac{d}{dt} i_{fe\beta} \]  \hspace{1cm} (6.35)

Figure 6.23 Stationary and synchronous reference frame
The stationary reference frame equations are transformed to synchronously rotating reference frame; the orientation being done with respect to the supply voltage space phasor as indicated earlier. The relative orientation of the stationary and synchronous reference frames is shown in figure 6.23. To maintain compatibility with the d-axis and q-axis definitions used in rotor side control, the supply voltage phasor axis is taken as the q-axis in this case. The unit vectors \( \cos \theta \) and \( \sin \theta \) can be directly obtained from \( u_{ac} \) and \( u_{ac\beta} \) respectively. In terms of the d-axis and q-axis variables eqns.(6.34) and (6.35) can be rewritten as follows:

\[
L_{fe} \frac{d}{dt} [(i_{eq} + j i_{fed}) e^{j \theta}] + \frac{d \theta}{dt} \left[ \frac{L_{fe}}{j} (i_{eq} + j i_{fed}) \right] e^{j \theta} + \frac{d}{dt} \left[ L_{fe} (i_{eq} + j i_{fed}) \right] e^{j \theta} = 0
\]

Eqn.(6.36) describes the system dynamics in the stationary coordinates in terms of the synchronous reference frame variables. Since the orientation is done with respect to the grid voltage, \( u_{acd} \) is zero. Transforming this to the rotating reference frame by multiplying both sides with \( e^{-j \theta} \) and separating the real and imaginary parts the following d-axis and q-axis equations can be obtained.

\[
u_{fed} + L_{fe} \frac{d}{dt} i_{fed} + \omega_s L_{fe} i_{eq} = 0
\]

(6.37)
$$u_{\text{eq}} + L_{\text{le}} \frac{d}{dt} i_{\text{eq}} - \omega_s L_{\text{le}} i_{\text{eq}} = u_{\text{acq}} \tag{6.38}$$

Due to the transformation, rotational emf terms appear in the voltage equations in the synchronous reference frame, giving rise to cross-coupling between the two axes. These rotational emf terms are required to be compensated by appropriate feed forward signals in order to independently control the active and reactive components of current.

![Figure 6.24 DC Bus Model](image)

The dynamics of the dc bus voltage can be modeled by considering the balance between the active power flow between the ac and the dc sides. If the series inductor is lossless,

$$\frac{2}{3} u_{\text{acq}} i_{\text{eq}} = u_{\text{acq}} i_{\text{dc}} \tag{6.39}$$

where $i_{\text{dc}}$ is the dc bus current as indicated in figure 6.24. This current can be written in terms of the capacitor charging current $i_c$ and, the load current $i_l$ as

$$i_{\text{dc}} = i_c + i_l = C \frac{du_{\text{dc}}}{dt} + i_l \tag{6.40}$$
Substituting $i_{dc}$ from equn.(6.40) in equn.(6.39) the following equation can be derived.

$$C \frac{du_{dc}}{dt} = \frac{2}{3} \frac{u_{acq}}{u_{dc}} i_{eq} - i_i$$  \hspace{1cm} (6.41)

Since the dc bus voltage is regulated within a narrow band and, the ac side voltage is nominally constant, the factor $\left( \frac{u_{acq}}{u_{dc}} \right)$ may be considered as a constant ratio to transform the synchronous reference frame current to the dc link current.

The model is schematically shown in figure 6.24.

6.10 Front end converter controller design

6.10.1 Design of the current controller

The design of the current controller is similar to that discussed in rotor side control. Two independent controllers are used for controlling the d-axis and q-axis currents. For the sake of completeness design of proportional controller and proportional-integral controller are discussed in detail.

a) Proportional Controller

Let the desired current loop dynamics in the q-axis be given by,

$$T_{ife} \frac{di_{eq}}{dt} + i_{eq} = \frac{i^*_{eq}}{K_{ife}}$$ \hspace{1cm} (6.42)

where $T_{ife}$ is the desired current-loop time constant

and, $K_{ife}$ is the current sensor gain.
Substituting \( \frac{d i_{feq}}{dt} \) from eqn.(6.42) in eqn.(6.38) gives the q-axis component of the instantaneous inverter terminal voltage required to produce the desired current dynamics.

\[
u_{feq} = u_{acq} + \omega_s L_{fe} i_{fed} - \frac{L_{fe}}{T_{fe} K_{fe}} \left( i_{feq}^* - K_{ife} i_{feq} \right)
\]

(6.43)

The inverter can be modeled as a constant gain block \( G_{fe} \) as explained in section 6.4.2(a). The reference for the q-axis component of rotor voltage is, therefore, given by

\[
u_{feq}^* = \frac{u_{feq}}{G_{fe}}
\]

\[
= \frac{u_{uqf}}{G_{fe}} + \omega_s L_{fe} i_{fed} - \frac{L_{fe}}{T_{ife} K_{ife} G_{fe}} \left( i_{feq}^* - K_{ife} i_{feq} \right)
\]

(6.44)

The plant along with the controller for the q-axis is shown in figure 6.25. The current sensor gain \( K_{ife} \) is taken as unity to simplify the diagram.

Assuming same current loop dynamics for the d-axis, the reference for the d-axis component of the rotor voltage can be expressed as the following:

\[
u_{fed}^* = \frac{u_{fed}}{G_{fe}}
\]

\[
= - \frac{\omega_s L_{fe} i_{eq}}{G_{fe}} - \frac{L_{fe}}{T_{ife} K_{ife} G_{fe}} \left( i_{fed}^* - K_{ife} i_{fed} \right)
\]

(6.45)
If the impressed inverter voltages are in accordance with eqn. (6.44), the active and reactive components of the ac side current can be controlled independently.

![Block diagram of q-axis proportional front end current controller](image)

**Figure 6.25 Block diagram of q-axis proportional front end current controller**

So far it has been assumed that the resistive drop for the inductor is negligible, which in practice is a valid assumption. However, if a proportional controller is used, the inclusion of the resistive drop as compensating terms in (6.44) and (6.45) minimizes the steady-state error.
b) Proportional Integral Controller

The design of the proportional-integral controller is similar to that discussed in section 6.4.2(b). The resistive drop is taken into consideration and the plant is represented by a first-order lag along with the cross-coupling and input voltage terms. The q-axis plant and controller are given in figure 6.26. The PI time-constant $T_{pifc}$ is made equal to $T_{fe} (= L_{fe} / R_{fe})$ and the proportional gain $K_{pifc}$ is selected as $\left( \frac{T_{fe}}{T_{ifc}} \right) R_{fe}$, where $T_{ifc}$ is the effective current-loop time constant.

![Figure 6.26 Block diagram of q-axis proportional-integral front end current controller](image-url)

Figure 6.26 Block diagram of q-axis proportional-integral front end current controller
6.10.2 Design of the voltage controller

Since the primary objective of the controller is to regulate the dc bus voltage within a narrow band, a proportional integral controller is the obvious choice. It may be appreciated that the response of the voltage controller need not be very fast. It is, however, desirable that the transient undershoot or overshoot in the dc bus voltage due to sudden variations of load in the dc side is limited to a minimum, normally within 5%. This is achieved by using the dc-side load current as a feed-forward term to the voltage controller. The structure of the controller, along with the plant is shown in figure 6.27(a). It is assumed that the current control loop is much faster than the outer voltage loop. So the dynamics of the current loop can be neglected while designing the voltage controller. The forward path transfer function of the system becomes,

\[
\frac{K_{pvfe} \left( 1 + s T_{vfe} \right)}{CT_{vfe} s^2}
\]

From the bode plot of the system, shown in figure 6.27(b), it can be inferred that \(1/T_{vfe}\) has to be selected slightly lower than the desired band width. The gain \(K_{pvfe}\) is then adjusted to make the phase margin close to 90°.

![Figure 6.27(a) Structure of PI controller](image_url)
6.11 Simulation Results – Front end Converter

The SIMULINK model of the front end converter power circuit is shown in figure 6.28. The three phase inverter block generates the inverter terminal voltages taking the gating signals as its input. The ac side dynamic equations in the stationary reference frame i.e. equn.(6.30) through equn.(6.31) are modeled in the subsequent block. The output from this block are the inductor currents, which are then multiplied with the switch status to form the dc side current in the demodulator block. Finally, the dynamics of the capacitor voltage, given by equn.(6.41), is modeled to get the dc bus voltage.

The modeling of the entire system along with the feedbacks and controllers is given in figure 6.29. The individual functional modules can be identified clearly from this diagram.
Figure 6.28 SIMULINK model of the three phase front end converter power circuit

Figure 6.29 SIMULINK system model of the three phase front end converter

The simulation results for the three phase front end converter under various transient conditions are given in figure 6.30 through figure 6.34. The parameters for the power circuit and the controller used in this simulation are listed in a
MATLAB file in Appendix B. The controller module is modeled in per unit in order to emulate the actual implementation of the controller.

In figure 6.30(a), responses of $i_{eq}$ and $i_{fed}$ for step change in $i^*_{eq}$ from 0 to 0.5 p.u. with $i^*_{fed} = 0$, are shown. The designed current loop time constant is 2ms. The corresponding response of the ac side current is given in figure 6.30 along with the supply voltage. Since $i^*_{fed} = 0$, the input power factor is observed to be unity. The reversal of active current from 0.5 p.u. to -0.5 p.u. under unity power factor operation is shown in figure 6.31. Current responses for step change in $i^*_{fed}$ from 0 to 0.25 p.u. with $i^*_{eq}$=0.5 p.u. are plotted in figure 6.32. The input current waveform shows that the front end operates at a leading power factor, thereby supplying reactive power to the source. There is, of course, a limit to the reactive current that can be injected. This depends on the dc bus voltage and the line side inductance value. Since the reactive voltage drop adds to the line voltage in the same phase, the maximum value of $i_{fed}$ can be written by the following equation.

$$i_{fed, \text{max}} = \sqrt{2} \frac{3}{2} \frac{\frac{u_{dc}}{2\sqrt{2}m_{\text{max}} - u_s}}{\omega L}$$  \hspace{0.5cm} (6.47)

The decoupling of the dynamics of the d-axis and q-axis current loops is clearly evident from these plots.

The initial charging of the dc bus voltage and, transient due to application of positive load is shown in figure 6.33. The response of the dc bus voltage when
the load is reversed is plotted in figure 6.34. The voltage loop time constant is assumed to be 100ms.

Figure 6.30(a) Response of $i_{feq}$ and $i_{fed}$

Figure 6.30(b) $u_{ac}$ and response of $i_{fe}$
Figure 6.31(a) Response of $i_{feq}$ and $i_{fed}$

Figure 6.31(b) $u_{ac}$ and response of $i_{fe}$
Figure 6.32(a) Response of $i_{feq}$ and $i_{fed}$

Figure 6.32(b) $u_s$ and response of $i_{fe}$
Figure 6.32 Simulation results for step change in $i^*_{fed}$ from 0 to 0.25 p.u. with $i^*_{feq} = 0.5$ p.u.
Figure 6.33(a) Response of $u_{dc}$

Figure 6.33(b) Response of $i_{feq}$

Figure 6.33 Simulation results of response of dc bus voltage when a positive load of 0.75 p.u. is suddenly applied on the dc side
Figure 6.34(a) Response of $u_{dc}$

Figure 6.34(b) Response of $i_{feq}$

Figure 6.34 Simulation results of response of dc bus voltage when the dc side load is reversed from 0.35 p.u. to -0.35 p.u.
6.12 Conclusion

A stator flux oriented model has been derived for the wound rotor induction machine. Current controllers designed in the field reference frame comprise proportional or proportional-integral controllers with subsequent addition or subtraction of the compensating terms. The design method is simple as it directly follows from the rotor voltage equations. Simulation results show that the dynamics of the active and reactive current loops are decoupled as required. The front end converter is modeled in the stator voltage reference frame. The structure of the current loop is similar to that for the rotor side control. The front end converter has been exhaustively simulated for forward and reverse power flow conditions. It is shown that leading power factor operation is possible upto a certain limit. The voltage controller exhibits excellent transient response during sudden impact of load on the dc bus.