A Robust Adaptive Sliding-Mode Controller for Slip Power Recovery Induction Machine Drives

J. Soltani and A. Farrokh Payam

Abstract—In this paper a robust nonlinear controller is presented for Doubly-Fed Induction Machine (DFIM) drives. The nonlinear controller is designed based on combination of Sliding-Mode (SM) and Adaptive-Backstepping control techniques. Using the fifth order model of DFIM in a stator d, q axis reference frames with stator currents and rotor flux components as state variables, First a SM controller is designed in order to follow a linear reference model. Then, the SM control and adaptive backstepping control approach are combined to derive a robust composite nonlinear controller for DFIM, which makes the drive system robust and stable against the parameters uncertainties and external load torque disturbance. In this drive system two back-to-back voltage source two level SVM-PWM inverters are employed in the rotor circuit, to make the drive system capable of operating in motoring and generating modes below and above synchronous speed. Computer simulation results obtained, confirm the effectiveness and validity of the proposed control approach.

Index Terms—Doubly fed induction machine, nonlinear sliding-mode, adaptive backstepping, torque, flux control.

I. INTRODUCTION

So far, the vector control (VC) and Direct torque control (DTC) methods have been applied to the three phase squirrel cage induction machine drives [1]. Although the field oriented control method in the stator or magnetizing reference frame are applied to the DFIM drives [2], [3], little attention has been given to the DTC and flux control of these types of drives.

In field oriented methods applied to DFIM drives, the voltage drop across the stator leakage impedance is neglected. Such an assumption, forces a steady-state error both in motoring and generating modes of operation. In [4] the bang-bang DTC control method is combined with direct rotor flux field oriented control method and is applied to an adjustable doubly fed induction motor drive. In [4], the controller is designed based on neglecting the voltage drop across the rotor resistance. In [5], [6], a backstepping tracking controller has been introduced for a DFIM drive. The control method of [5], [6] is used only in generating mode of operation upon unity of power factor, measured on the stator side of the machine. This paper describes a nonlinear controller for DFIM drives based on combination of SM control and adaptive backstepping control approaches. The SM control forces the system state nominal trajectories to follow a linear reference model.

II. DFIM MODEL

The electrical configuration of a DFIM drive system is depicted in Fig. 1. This system consists of a three-phase wound rotor induction machine connected through a back-to-back converter scheme, which in turn is connected to the main ac supply. The drive system shown in Fig. 1 can operate as a motor or a generator below and above the synchronous speed. The different modes of operation of this drive system are shown in Table I [8].

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Under assumption of linear magnetic circuits and balanced operating condition, the equivalent two-phase model of the symmetrical DFIM with stator connected to line, represented in fixed stator d-q reference frame is

\[
\frac{di_d}{dt} = -\frac{R}{L}i_d + \frac{1}{L}i_q + \omega L_\alpha \psi_{d,\alpha} + \frac{\omega L}{L} \psi_{d,\alpha} + \frac{u_{d,\alpha}}{L} - \omega L\psi_{d,\alpha} + \frac{u_{d,\alpha}}{L} - \omega L\psi_{d,\alpha} + \frac{u_{d,\alpha}}{L} \\
\frac{di_q}{dt} = -\frac{R}{L}i_q + \frac{1}{L}i_d + \omega L_\alpha \psi_{q,\alpha} + \frac{\omega L}{L} \psi_{q,\alpha} + \frac{u_{q,\alpha}}{L} - \omega L\psi_{q,\alpha} + \frac{u_{q,\alpha}}{L} - \omega L\psi_{q,\alpha} + \frac{u_{q,\alpha}}{L} \\
(1)
\]

where \(i_d, i_q, u_d, u_q, R, \) and \(L\) denote stator current, rotor flux linkage, stator terminal voltage, rotor terminal voltage, resistance and inductance, respectively. The subscripts s and r stand for stator and rotor while subscripts d and q stand for vector component with respect to a fixed stator reference frame. \(\omega_0\) denotes the rotor electrical angular speed and \(L_m\) is mutual inductance. \(L_{\alpha} = L_{11} - (1 - L_{22} / L_{22})\) is the redefined leakage inductance.

The DFIM generated torque can be expressed in terms of stator currents and rotor flux linkage as follows

\[
T_e = \frac{3P}{2} L_{\alpha}(\psi_{d,\alpha} i_q - \psi_{q,\alpha} i_d) \\
(2)
\]

where \(P\) is number of poles.

The mechanical equation is given by

\[
J \frac{d\omega_m}{dt} + B \omega_m + T_L = T_e \\
(3)
\]

where \(J\) and \(B\) denote the moment of inertia and friction coefficient, respectively, \(T_e\) is the external load torque and \(\omega_m\) is the rotor mechanical angular speed \((\omega_m = (P/2)\omega_0)\).

Let

\[
x = \begin{bmatrix} i_d & i_q & \psi_{d,\alpha} & \psi_{q,\alpha} \end{bmatrix}^T \\
(4)
\]

be the state vector and for generated torque \(T_e\) defined by

\[
y = T_e = \frac{3P}{2} L_{\alpha}(\psi_{d,\alpha} i_q - \psi_{q,\alpha} i_d) \\
(5)
\]

III. NONLINEAR SLIDING MODE CONTROL

Equation (1) in compact form rewritten as

\[
\dot{x} = f(x) + g_1 u_d + g_2 u_q \\
(6)
\]

where \(x\) is defined in (4) and

\[
f(x) = \begin{bmatrix}
\frac{R}{L} i_d + \frac{1}{L} i_q + \omega L_\alpha \psi_{d,\alpha} + \frac{\omega L}{L} \psi_{d,\alpha} + \frac{u_{d,\alpha}}{L} - \omega L\psi_{d,\alpha} + \frac{u_{d,\alpha}}{L} - \omega L\psi_{d,\alpha} + \frac{u_{d,\alpha}}{L} \\
\frac{R}{L} i_q + \frac{1}{L} i_d + \omega L_\alpha \psi_{q,\alpha} + \frac{\omega L}{L} \psi_{q,\alpha} + \frac{u_{q,\alpha}}{L} - \omega L\psi_{q,\alpha} + \frac{u_{q,\alpha}}{L} - \omega L\psi_{q,\alpha} + \frac{u_{q,\alpha}}{L}
\end{bmatrix}
\]

\[
g_1 = \begin{bmatrix}
-\frac{L_{\alpha}}{L} & 0 & 1 & 0
\end{bmatrix}^T \\
(8)
\]

\[
g_2 = \begin{bmatrix}
0 & -\frac{L_{\alpha}}{L} & 0 & 1
\end{bmatrix}^T \\
(8)
\]

Assuming the machine torque \(T_r\) and the norm of the rotor flux modulus \(|\psi_r| = \psi_{r,\alpha}^2 + \psi_{r,\alpha}^2\) are to be the system control outputs. Based on input-output feedback linearization

\[
h_1(x) = \frac{3P}{2} L_{\alpha}(\psi_{d,\alpha} i_q - \psi_{q,\alpha} i_d) \\
(9)
\]

\[
h_2(x) = \psi_{r,\alpha}^2 + \psi_{r,\alpha}^2 \\
(9)
\]

Introducing the new variables as

\[
z_1 = h_1(x) \\
z_2 = h_2(x) \\
(10)
\]

Then, the dynamic model of DFIM in new coordinates is given by

\[
\begin{bmatrix}
\dot{z}_1 \\
\dot{z}_2
\end{bmatrix} = \begin{bmatrix}
L_{\alpha} h_2 & L_{\alpha} h_1 & L_{\alpha} h_1 & L_{\alpha} h_1
\end{bmatrix} \begin{bmatrix}
u_{d,\alpha} \\
u_{q,\alpha}
\end{bmatrix} \\
(11)
\]

where

\[
L_{\alpha} h_2 = 2 \frac{R}{L} \left(i_{d,\alpha} \psi_{r,\alpha} - i_{q,\alpha} \psi_{r,\alpha} \right) - 2 \frac{R}{L} \left(\psi_{r,\alpha}^2 + \psi_{r,\alpha}^2 \right) \\
L_{\alpha} h_1 = \frac{3P}{2} \frac{L_{\alpha}}{L} \times \left(-\left(\frac{R}{L} + \frac{R}{L} + \frac{R}{L}\right) (i_{d,\alpha} \psi_{r,\alpha} - i_{q,\alpha} \psi_{r,\alpha} \right) - \omega L_{\alpha} \left(\psi_{r,\alpha}^2 + \psi_{r,\alpha}^2 \right) - \omega L_{\alpha} \left(\psi_{r,\alpha}^2 + \psi_{r,\alpha}^2 \right) + u_{d,\alpha} \psi_{r,\alpha} - u_{q,\alpha} \psi_{r,\alpha} \right)
\]

\[
L_{\alpha} h_1 = 2 \psi_{r,\alpha} \\
L_{\alpha} h_2 = 2 \psi_{r,\alpha} \\
(12)
\]

\[
L_{\alpha} h_1 = -\frac{3P}{2} \frac{L_{\alpha}}{L} (i_{d,\alpha} + \frac{L_{\alpha}}{L} \psi_{r,\alpha}) \\
L_{\alpha} h_2 = -\frac{3P}{2} \frac{L_{\alpha}}{L} (i_{q,\alpha} + \frac{L_{\alpha}}{L} \psi_{r,\alpha}) \\
(12)
\]

Furthermore, a nonlinear state feedback control inputs is employed as

\[
\begin{bmatrix}
\dot{u}_{d,\alpha} \\
\dot{u}_{q,\alpha}
\end{bmatrix} = \begin{bmatrix}
L_{\alpha} h_2(x) & L_{\alpha} h_1(x) \\
L_{\alpha} h_1(x) & L_{\alpha} h_2(x)
\end{bmatrix} \begin{bmatrix}
u_{d,\alpha} \\
u_{q,\alpha}
\end{bmatrix} \\
(13)
\]

Then, the dynamic system (11) becomes

\[
\begin{bmatrix}
\dot{z}_1 \\
\dot{z}_2
\end{bmatrix} = \begin{bmatrix}
L_{\alpha} h_2(x) & 1 & 0 \\
L_{\alpha} h_1(x) & 0 & 1
\end{bmatrix} \begin{bmatrix}
u_{d,\alpha} \\
u_{q,\alpha}
\end{bmatrix} \\
(14)
\]

From (14) the reference model is introduced as

\[
\begin{bmatrix}
\dot{z}_{1,\alpha} \\
\dot{z}_{2,\alpha}
\end{bmatrix} = \begin{bmatrix}
-a_{m1} & 0 & a_{m2} & 0 & a_{m3} \psi_{r,\alpha}^2 & a_{m4} \\
0 & -a_{m2} & a_{m1} & 0 & \psi_{r,\alpha}^2
\end{bmatrix} \\
(15)
\]

where \(a_{m1}, a_{m2}, a_{m3}\) and \(a_{m4}\) are the positive constants. Using (14) and (15) the system error dynamics is obtained as
\[ \dot{e}_z = A(x) + B(x)\dot{U} \]  
where
\[ e_z = [z_1 - z_{m1}, z_2 - z_{m2}]^T = [e_{z1}, e_{z2}]^T \]  
with
\[ A(x) = \begin{bmatrix} L_i h_i(x) \\ L_f h_f(x) \end{bmatrix}, \quad B(x) = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \]  
\[ \dot{U} = \begin{bmatrix} \pi_0 \\ \pi_\psi \end{bmatrix} = \begin{bmatrix} \dot{u}_\psi + a_m z_{m1} - a_m \psi \sigma_i \\ \dot{u}_\sigma + a_m z_{m2} - a_m \psi \sigma_i \end{bmatrix} \]  

According to the system shown in (16), the sliding switching surfaces are chosen as
\[ \sigma(e_z) = Se_z(x) \]  \tag{19} 
where \( S \in \mathbb{R}^{2x2} \) is a constant linear matrix so that the inverse of \( SB(x) \) must exist for all \( x \), i.e., \( \det(SB(x)) \neq 0 \) for all \( x \). Combining (16) and (19), gives
\[ \sigma = Se_z = SA(x) + SB\dot{U} = -Q\text{sgn}(\sigma) - K\sigma \]  \tag{20} 
where
\[ Q = \begin{bmatrix} q_1 & 0 \\ 0 & q_2 \end{bmatrix}, \quad K = \begin{bmatrix} k_1 & 0 \\ 0 & k_2 \end{bmatrix}, \quad q_i, k_i > 0, i = 1, 2 \]  \tag{21} 
where
\[ \text{sgn}(\sigma) = \begin{cases} 1 & \text{as } \sigma > 0 \\ -1 & \text{as } \sigma < 0 \end{cases} \]  \tag{22} 

From (20) based on Lyapunov theory, the SM controller is obtained as
\[ \dot{U} = -(SB)^{-1}[SA(x) + Q\text{sgn}(\sigma) + K\sigma] \]  \tag{23} 

Using the control low of (23), the reachability of SM control is guaranteed.

**Theorem 1:** With the developed nonlinear sliding-mode controller (23) and a stable sliding surface (19), the reaching condition \( \sigma^r \sigma < 0 \) is satisfied, and the controlled system (16) will be stabilized.

**Proof:** From the designed dynamic sliding surface (22), the following equation can be derived
\[ \dot{\sigma}_z = -q_i \text{sgn}(\sigma_i) - k_i \sigma_i, i = 1, 2 \]  \tag{24} 

Multiplying \( \sigma_i \) on the both side of the above equation yields
\[ \sigma_i \dot{\sigma}_i = -q_i \sigma_i \text{sgn}(\sigma_i) - k_i \sigma_i^2 = -q_i |\sigma_i| k_i |\sigma_i|^2 < 0 \]  \tag{25} 

From the above analysis, it is evident that the reaching condition is guaranteed.

**Remark 1:** Using the model-following nonlinear SM control based on the state-coordinate transformed model for the DFIM, the transient of the generated torque and rotor flux amplitude can be regulated through a linear reference model.

In practical and industrial applications, the rotor and stator resistance \( R_s \) and \( R_r \) vary in temperature and the magnetic flux saturation. Therefore, nonlinear SM controller that makes the DFIM stable and robust against the parameters variations is proposed in the next section.

### IV. Adaptive Backstepping Sliding-Mode Control

Adaptive backstepping control has the features of having more than one estimate per unknown parameter for controlled system with mismatched uncertainties [9].

When the system parameters deviate from the nominal values, especially the resistances \( R_s \) and \( R_r \), the tracking error model (17) can be rewritten as
\[ \dot{e}_{z1} = L_i h_i(x) + \bar{u}_{o_r} + \phi d_1(x) \]  
\[ \dot{e}_{z2} = L_f h_f(x) + \bar{u}_{o_q} + \phi d_2(x) \]  
\[ \dot{e}_z = [A(x) + \Delta A(x)] + B(x)\dot{U} \]  \tag{26} 

where \( \phi_i \), \( i = 1, 2 \) and \( \Delta A(x) \) denote in fact uncertainties defined by
\[
\begin{align*}
\dot{\alpha} &= \frac{R_s L_m}{L_f} L_r L_m + \frac{\Delta R_s L_m}{L_f} L_r L_m \\
\dot{\beta} &= \frac{\Delta R_s L_m}{L_f} L_r L_m
\end{align*}
\]  \tag{27} 

Therefore,
\[
\begin{align*}
L_\omega h_i(x) &= \frac{2\Delta R_s L_m}{L_f} \left[ i_w \psi_{i_w} + \psi_{i_w} \right] \\
L_\omega h_f(x) &= \frac{3\Delta R_s L_m}{2L_f} \left[ \Delta R_s L_m \left(i_w \psi_{i_w} + \psi_{i_w}ight) - \Delta \alpha \left(i_w \psi_{i_w} - \psi_{i_w}\right) \right] \\
\phi d_1(x) &= \frac{2\Delta R_s L_m}{L_f} \left[ i_w \psi_{i_w} + \psi_{i_w} \right] \\
\phi d_2(x) &= \frac{3\Delta R_s L_m}{2L_f} \left[ \Delta R_s L_m + \Delta \alpha \right] \left(i_w \psi_{i_w} - \psi_{i_w}\right)
\end{align*}
\]  \tag{28} 

that is \( \left[ \phi d_1(x), \phi d_2(x) \right]^T = \Delta A(x) \).

Since \( R_s \) and \( R_r \) are sensitively varied with thermal drift, we assume that \( |\phi_i| \) \( (i = 1, 2) \) is the unknown and bounded constant.

Deriving the system errors \( e_z \) with respect to time \( t \), yields
\[
\begin{align*}
\dot{e}_{z1} &= L_i h_i(x) + \bar{u}_{o_r} + \phi d_1(x) - (\phi - \phi_k) d_1(x) \\
\dot{e}_{z2} &= L_f h_f(x) + \bar{u}_{o_q} + \phi d_2(x) - (\phi - \phi_k) d_2(x)
\end{align*}
\]  \tag{29} 

where \( \phi_k \) \( (i = 1, 2) \) is the estimate of \( \phi, k_i, i = 1, 2 \) is a positive constant feedback gain.

It is obvious that the controllers \( \bar{u}_{o_r} \) and \( \bar{u}_{o_q} \) are decoupling with respect to two dynamic models, \( [e_{z1}, e_{z2}] \). From (29), the \( d \) axes SM control is designed as
\[ \bar{u}_{o_r} = -L_i h_i(x) - \hat{\phi} d_1(x) - k_1 e_{z1} - \rho_1 \text{sgn}(e_{z1}) \]  \tag{30} 

where \( k_1 > 0 \) and \( \rho_1 \) is chosen as follows
\[ \rho_1 d_1(x) \geq |\phi - \phi_k| d_1(x) \]
when \( \hat{\phi} \) is in transient and the adaptation law of \( \hat{\phi} \) is given by
\[ \dot{\phi}_1 = g_1 e_{21} d_1(x) \]  
where \( g_1 > 0 \) is the adaptation gain. Similarly, the \( q \) axes SM control is designed as

\[ \tau_{\phi} = L f h(x) - \dot{\phi}_2 d_2(x) - k_2 e_{22} - \rho_2 \text{sgn}(e_{22}) \]  
where \( k_2 \) is positive constant feedback gain and

\[ \hat{\phi}_2 = g_2 e_{22} d_2(x) \]  
where \( g_2 > 0 \) is the adaptation gain and \( \rho_2 \) is chosen as follows

\[ \rho_2 d_2(x) \geq \left| \hat{\phi}_2 - \phi_2 \right| d_2(x) \]

when \( \phi_2 \) is in transient.

**Theorem 2:** Using the controller described by (30)-(33), the torque and flux amplitude DFIM is stable and robust subject to the parameters mismatched uncertainties.

Proof: Defining the following Lyapunov function

\[ V_1 = \frac{1}{2} e_{21}^2 + e_{22}^2 + \frac{1}{g_1} \left( \phi_1 - \hat{\phi}_1 \right)^2 + \frac{1}{g_2} \left( \hat{\phi}_2 - \phi_2 \right)^2 \]  
(34)

Derivative (34) with respect to time \( t \) and combining with (29), gives

\[ \dot{V}_1 = e_{21} \left[ L h_1(x) + \tau_{\phi} + \dot{\phi}_1 d_1(x) - (\hat{\phi}_1 - \phi_1) d_1(x) \right] + e_{22} \left[ L h_1(x) + \tau_{\phi} + \dot{\phi}_2 d_2(x) - (\hat{\phi}_2 - \phi_2) d_2(x) \right] + \frac{1}{g_1} (\phi_1 - \hat{\phi}_1) \dot{\phi}_1 + \frac{1}{g_2} (\hat{\phi}_2 - \phi_2) \dot{\phi}_2 \]  
(35)

Substituting the control laws (30) and (32) and the adaptation laws (31) and (33) into (35), the following inequality can be deduced

\[ \dot{V}_1 = -k_1 e_{21}^2 - k_2 e_{22}^2 \leq 0 \]  
(36)

Defining the following equation

\[ M(t) = k_1 e_{21}^2 + k_2 e_{22}^2 \geq 0 \]  
(37)

Furthermore,

\[ V_1(t) = V_1(e(0), \dot{\phi}(0)) + \int_0^t \dot{V}_1(\tau) d\tau \]  
(38)

\[ = V_1(e(0), \dot{\phi}(0)) - \int_0^t M(\tau) d\tau \]

where \( e = [e_{21}, e_{22}]^T \) and \( \dot{\phi} = [\dot{\phi}_1, \dot{\phi}_2]^T \). From the definition of the Lyapunov function \( V_1(t) \geq 0 \) and the above equation, one can obtain that

\[ \lim_{t \to \infty} M(t) d\tau \leq V_1(e(0), \dot{\phi}(0)) < \infty \]  
(39)

Based on the Barbalat’s Lemma [10], we can obtain

\[ M(t) \to 0 \text{ as } t \to \infty \]  
(40)

That is, \( e_{21} \) and \( e_{22} \) will converge to zero as \( t \to \infty \). Therefore, the proposed controller is stable and robust, even if parameters uncertainties exist.

V. STABILIZATION OF ROTOR DC-LINK VOLTAGE

Using the method described in [9], the \( d \) and \( q \) axis current equations corresponding to main ac power supply are given by

\[ \frac{di_d}{dt} = \frac{1}{L} (v_d - R i_d + \omega_d L i_q - v_{di}) \]  
(41)

\[ \frac{di_q}{dt} = \frac{1}{L} (-R i_q - \omega_d L i_d - v_{qi}) \]

One may note that in this reference frames the \( d \) axis is in coincided with the main space voltage vector.

Considering \( i_d' \), \( i_q' \) as reference currents, therefore

\[ e_i = i_d - i_d' \]

\[ e_q = i_q - i_q' \]

as result, the system error dynamic is

\[ \dot{e}_d = \frac{1}{L} R i_d + \alpha_d i_q - \frac{v_{di}}{L} \]

\[ \dot{e}_q = -\frac{1}{L} R i_q - \omega_d L i_d - \frac{v_{qi}}{L} \]

Defining the ac side inverter reference voltages as

\[ v_{di} = L (\frac{v_d}{L} - \frac{R}{L} i_d + \omega_d i_q - i_d' - k e_i) \]  
(44)

\[ v_{qi} = L (\frac{v_q}{L} - \frac{R}{L} i_q - \omega_d i_d - i_q' - k e_q) \]

Linking (43) and (44), gives

\[ \dot{e}_q = -k e_i \]

\[ \dot{e}_q = -k e_q \]

Considering a Lyapunov function as

\[ V = \frac{1}{2} e_i^2 + \frac{1}{2} e_q^2 \]  
(46)

therefore

\[ \dot{V} = \dot{e}_i e_i + \dot{e}_q e_q \leq -k (e_i^2 + e_q^2) < 0 \]  
(47)

Based on above control strategy the block diagram of rotor dc link voltage controller is depicted in Fig. 2.

VI. ACTIVE AND REACTIVE STATOR POWER CONTROL IN GENERATING MODE

For electric energy generation, it is usually required to regulate the stator active-reactive power, whose references are assumed to be \( P_s' \) and \( Q_s' \) respectively. These quantities in a special two axis rotating reference frame which is defined in the previous section are given by [11]

\[ i_d' = \frac{Q_s'}{2 U} \]

\[ i_q' = \frac{P_s'}{2 U} \]  
(48)

Therefore, flux references calculated from

\[ \psi_d' = \frac{1}{\beta_b h_b} (\frac{R}{\sigma} i_d' - \omega_d i_q') \]  
(49)

\[ \psi_q' = \frac{1}{\beta_b h_b} (\frac{R}{\sigma} i_q' - \omega_d i_d' - \frac{1}{\sigma}) \]

so rotor flux reference and torque reference are calculated as below.


\[ \psi_r' = \sqrt{\psi_d'^2 + \psi_q'^2} \]

\[ T_r' = \mu(\psi_d'i_q' - \psi_q'i_d') \]  

(50)

VII. SYSTEM SIMULATION

The overall block diagram of the proposed control approach is shown in Fig. 3. A C++ computer program was developed to model this system on P.C.

In this program, a static Runge-Kutta fourth order method is used to solve the system equations. The effectiveness and validity of the proposed approach is tested for a three-phase 5 kW, 380 V, six poles, 50 Hz DFIM drive [11] by simulation.

Simulation results shown in Fig. 4 are obtained for the system motoring mode of operation below and above synchronous speed. These results are obtained in the condition of an exponential speed reference from 0 to 260 (rad/sec) rise up to 350 (rad/sec) at t=2 sec with \( \tau_{00} = .07 \text{ sec} \), and reference flux signal that is construct from active and reactive power is shown in Fig. 4, a load torque reference profile which is also shown in Fig. 4 and \( R_n = R_{n+} = R_{n-} \).

Fig. 5 shows the system performance in the motoring mode of operation below and above synchronous speed. These results are obtained for the same condition described for Fig. 4 but for a \( R_n = 2R_{n+}, R_n = 2R_{n-} \).

Also in the all mode of operations, the torque and rotor reference flux signals are obtained based on desired active and reactive power reference profiles which are injected to the stator from main ac supply.

Fig. 6 shows the drive system performance in the generation mode of operation above synchronous speed. These results are obtained for \( R_n = 2R_{n+}, R_n = 2R_{n-} \) and torque reference profile and reference flux signal shown in Fig. 6 and a constant \( \omega_0 = 375 \) (rad/sec).

Fig. 7 shows the drive system performance in the generation mode of operation below synchronous speed.
These results are obtained for the same condition described for Fig. 6 and a constant $\omega_r = 280$ (rad/sec). One may note that in Figs. 6 and 7, the average active power injected to rotor circuit is positive. That is because of high rotor copper losses obtained by rotor resistance of $R_r = 2R_m$. Also in the above simulation results, $T_e$ is the machine generated torque, $\omega_r$ is the rotor speed in electrical (rad/sec), $\psi_r$ is the rotor flux linkage, $P_r$ is the power injected into the rotor circuit, $E$ is the rotor dc link voltage, $P_r$ and $Q_r$ are active and reactive power injected to the stator from ac supply.

**VIII. CONCLUSION**

In this paper a robust nonlinear controller has been designed for DFIM drives. The proposed controller is derived based on combination of SM control and adaptive backstepping control approach. The proposed composite control forces the system states trajectories to follow the nominal states which are obtained by a desired reference model, in spite of stator and rotor resistance uncertainties. The nonlinear control method has been tested for motoring and generating mode of operation below and above synchronous speed using two back-to-back two level SVM-PWM voltage sources inverters in rotor circuit. Furthermore the rotor dc link voltage is maintained constant also based on input-output control method, using a rotating synchronous reference frame with d axis coincide with the direction of space voltage vector of the main ac supply. Computer simulation results obtained, confirm the validity and capability of the proposed control approach.
REFERENCES


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