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An Adaptive Input-Output Feedback Linearization Controller for Doubly-Fed Induction Machine Drives

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Abstract: In this paper a nonlinear controller is presented for Doubly-Fed Induction Machine (DFIM) drives. The nonlinear controller is designed based on the adaptive input-output feedback linearization control technique, using the fifth order model of induction machine in fixed stator *d*, *q* axis reference frames with stator currents and rotor flux components as state variables. The nonlinear controller can perfectly track the torque and flux reference signals in spite of stator and rotor resistance variations. Two level SVM-PWM back-to-back voltage source inverters are employed in the rotor circuit, in order to make the drive system capable of operating in the motoring and generating modes below and above the synchronous speed. Computer simulation results obtained, confirm the effectiveness and validity of the proposed control approach.

Keywords: DFIM, Adaptive Input-Output Controller, Torque, Flux.

1 Introduction

So far, the vector control (VC) and Direct torque control (DTC) methods have been applied to squirrel cage induction machine drives [1]. Although field oriented vector methods have been applied to DFIM drives [2, 3], little attention has been given to DTC and flux control of these types of drives. In field oriented methods applied to DFIM drives, in order to design the rotor current controllers, the voltage drop across the stator leakage impedance needs to be neglected [2-4]. That is necessary in order to control the injected active and reactive powers to the stator independently. Such an assumption generated a steady-state error for both the motoring and generating modes of operation. In [5] the conventional bang-bang DTC control method is combined with the direct rotor flux field oriented control method and is applied to an adjustable speed DFIM drive. In [5], the DTC controller is designed based on neglecting the voltage drop across the rotor resistance. In [6,7], a backstepping tracking controller has been introduced for a DFIM drive. The methods of [6] and [7] have been proposed

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only for operation upon unity of the power factor, measured on the stator supply voltage side.

This paper describes an adaptive nonlinear controller for DFIM drives based on the adaptive input-output feedback linearization approach, using a fifth order model in stator fixed *d*, *q* axis reference frames with stator current and rotor flux vector components assumed as the state variables. It will be shown that the controller proposed in this paper is capable of perfectly tracking control of active and reactive power reference commands, which are injected into the stator in spite of machine resistance variations and external load torque disturbance. One may note that when the machine operates as a motor the injected active power into the stator must be positive. The overall system stability was proved by Lyapanov's theory. Moreover, using two level back-to-back SVM-PWM voltage source inverters in the rotor circuit, the proposed control method can be applied for motoring and generating modes of operation below and above the synchronous speed. In addition, the rotor dc link voltage is maintained constant also based on input-output linearization, using a rotating reference frame with the d axis coinciding with the space voltage vector of the main ac supply.

2 Doubly Fed Induction Machine Model

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

$$\begin{aligned}
\frac{\mathrm{d}i_{ds}}{\mathrm{d}t} &= -\left(\frac{R_s}{L_\sigma}+\frac{R_r L_m^2}{L_r^2 L_\sigma}\right)i_{ds}+\frac{R_r L_m}{L_r^2 L_\sigma}\psi_{dr}+\frac{\omega_r L_m}{L_r L_\sigma}\psi_{qr}+\frac{u_{ds}}{L_\sigma}-\frac{L_m}{L_r L_\sigma}u_{dr}\\
\frac{\mathrm{d}i_{qs}}{\mathrm{d}t} &= -\left(\frac{R_s}{L_\sigma}+\frac{R_r L_m^2}{L_r^2 L_\sigma}\right)i_{qs}+\frac{R_r L_m}{L_r^2 L_\sigma}\psi_{qr}-\frac{\omega_r L_m}{L_r L_\sigma}\psi_{dr}+\frac{u_{qs}}{L_\sigma}-\frac{L_m}{L_r L_\sigma}u_{qr}\\
\frac{\mathrm{d}\psi_{dr}}{\mathrm{d}t} &= \frac{R_r L_m}{L_r}i_{ds}-\frac{R_r}{L_r}\psi_{dr}-\omega_r\psi_{qr}+u_{dr}\\
\frac{\mathrm{d}\psi_{qr}}{\mathrm{d}t} &= \frac{R_r L_m}{L_r}i_{qs}-\frac{R_r}{L_r}\psi_{qr}+\omega_r\psi_{dr}+u_{qr},
\end{aligned} \tag{1}$$

where i_s, ψ_r, u_s, u_r, R and L denote stator currents, 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 the vector component with respect to a fixed stator reference frame. ω_r denotes the rotor electrical speed and L_m is the mutual inductance. $L_\sigma = L_s\left(1 - L_m^2/(L_r L_s)\right)$ is the redefined leakage inductance.

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

$$T_e = \frac{3P}{2}\frac{L_m}{L_r}(\psi_{dr} i_{qs} - \psi_{qr} i_{ds}), \tag{2}$$

where P is the number of pole pairs. The mechanical dynamic equation is given by

$$J\frac{\mathrm{d}\omega_m}{\mathrm{d}t} + B\omega_m + T_L = T_e, \tag{3}$$

where J and B denote the moment of inertia of the motor and viscous friction coefficient, respectively, T_L is the external load and ω_m is the rotor mechanical speed ($\omega_r = (P/2)\omega_m$).

Let

$$\boldsymbol{x} = \begin{bmatrix} i_{ds} & i_{qs} & \psi_{dr} & \psi_{qr} \end{bmatrix}^{\mathrm{T}} \tag{4}$$

be the state vector and let the generated torque T_e be the output y of the dynamic system (1) ,that is

$$y = T_e = \frac{3P}{2}\frac{L_m}{L_r}(\psi_{dr} i_{qs} - \psi_{qr} i_{ds}). \tag{5}$$

It is well known that torque control is very important for high-performance motion control [9]. However from (1) and (5), we can note that the generated torque T_e of DFIM is a nonlinear output with respect to state variables x of the dynamic model (1), $i_{ds}, i_{qs}, \psi_{dr}$ and ψ_{qr}. Therefore, it is difficult to evaluate the torque response from (5) by the control input u_{dr} and u_{qr} designed for the model (1). So based on the dynamic model (1), the torque control of induction machines is a task for industrial and practical applications.

3 Adaptive Input-Output Feedback Control

For the proposed nonlinear adaptive input-output feedback linearization controller, the state coordinate transformation is applied. Therefore the state-coordinates transformed model from (1) can be rewritten in a compact form as

$$\dot{\boldsymbol{x}} = f(x) + g_1 u_{dr} + g_2 u_{qr}, \tag{6}$$

where $\boldsymbol{x}$ is defined in (4) and

$$f(x)=\begin{bmatrix} -\left(\frac{R_s}{L_\sigma}+\frac{R_r L_m^2}{L_r^2 L_\sigma}\right)i_{ds}+\frac{R_r L_m}{L_r^2 L_\sigma}\psi_{dr}+\frac{\omega_r L_m}{L_r L_\sigma}\psi_{qr}+\frac{u_{ds}}{L_\sigma} \\ -\left(\frac{R_s}{L_\sigma}+\frac{R_r L_m^2}{L_r^2 L_\sigma}\right)i_{qs}+\frac{R_r L_m}{L_r^2 L_\sigma}\psi_{qr}-\frac{\omega_r L_m}{L_r L_\sigma}\psi_{dr}+\frac{u_{qs}}{L_\sigma} \\ \frac{R_r L_m}{L_r}i_{ds}-\frac{R_r}{L_r}\psi_{dr}-\omega_r\psi_{qr} \\ \frac{R_r L_m}{L_r}i_{qs}-\frac{R_r}{L_r}\psi_{qr}+\omega_r\psi_{dr} \end{bmatrix} \tag{7}$$

and

$$\begin{aligned} \boldsymbol{g}_1 &= \begin{bmatrix} -\frac{L_m}{L_r L_\sigma} & 0 & 1 & 0 \end{bmatrix}^{\mathrm{T}}, \\ \boldsymbol{g}_2 &= \begin{bmatrix} 0 & -\frac{L_m}{L_r L_\sigma} & 0 & 1 \end{bmatrix}^{\mathrm{T}}. \end{aligned} \tag{8}$$

At this stage the generated torque T_e and the squared modules of the rotor flux linkage, $|\psi_r|^2=\psi_{dr}{}^2+\psi_{qr}{}^2$, are requested to be the controlled output. Therefore, by considering

$$\begin{aligned} h_1(x) &= \frac{3P}{2}\frac{L_m}{L_r}(\psi_{dr}i_{qs}-\psi_{qr}i_{ds}) \\ h_2(x) &= \psi_{dr}^2+\psi_{qr}^2 \end{aligned} \tag{9}$$

The following notation is used for the lie derivative of a function $h(x):\Re^n\to\Re$ along a vector field $f(x)=(f_1(x),\ldots,f_n(x))$

$$L_f h(x)=\sum_{i=1}^{n}\frac{\partial h}{\partial x_i}f_i(x) \tag{10}$$

Iteratively, we define $L_f^i h=L_f(L_f^{i-1}h)$.

If the change of coordinates is defined as

$$\begin{aligned} z_1 &= h_2(x) \\ z_2 &= h_1(x) \end{aligned} \tag{11}$$

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

$$\begin{bmatrix} \dot{z}_1 \\ \dot{z}_2 \end{bmatrix} = \begin{bmatrix} L_f h_2 \\ L_f h_1 \end{bmatrix} + \begin{bmatrix} L_{g1} h_2 & L_{g2} h_2 \\ L_{g1} h_1 & L_{g2} h_1 \end{bmatrix} \begin{bmatrix} u_{dr} \\ u_{qr} \end{bmatrix}, \tag{12}$$

where

$$\begin{aligned}
L_f h_2 &= 2\frac{R_r L_m}{L_r}(i_{ds}\psi_{qr} - i_{qs}\psi_{dr}) - 2\frac{R_r}{L_r}(\psi_{dr}^2 + \psi_{qr}^2) \\
L_f h_1 &= \frac{3P}{2}\frac{L_m}{L_r} \times \Bigg(-\left(\frac{R_s}{L_\sigma} + \frac{R_r L_m^2}{L_r^2 L_\sigma} + \frac{R_r}{L_r} \right)(i_{qs}\psi_{dr} - i_{ds}\psi_{qr}) - \\
&\quad - \frac{\omega_r L_m}{L_r L_\sigma}(\psi_{dr}^2 + \psi_{qr}^2) - \omega_r(\psi_{qr} i_{qs} + \psi_{dr} i_{ds}) + \frac{u_{qs}\psi_{dr}}{L_\sigma} - \frac{u_{ds}\psi_{qr}}{L_\sigma} \Bigg) \\
L_{g1} h_2 &= 2\psi_{dr} \\
L_{g2} h_2 &= 2\psi_{qr} \\
L_{g1} h_1 &= \frac{3P}{2}\frac{L_m}{L_r}\left(i_{qs} + \frac{L_m}{L_r L_\sigma}\psi_{qr} \right) \\
L_{g2} h_1 &= -\frac{3P}{2}\frac{L_m}{L_r}\left(i_{ds} + \frac{L_m}{L_r L_\sigma}\psi_{dr} \right).
\end{aligned} \tag{13}$$

Furthermore, a nonlinear state feedback decoupling of the control inputs method is employed. We constructed the new control inputs as follows

$$\begin{bmatrix} \hat{u}_{dr} \\ \hat{u}_{qr} \end{bmatrix} = \begin{bmatrix} L_{g1} h_2(x) u_{dr} + L_{g2} h_2(x) u_{qr} \\ L_{g1} h_1(x) u_{dr} + L_{g2} h_2(x) u_{qr} \end{bmatrix}. \tag{14}$$

Then, system (12) becomes

$$\begin{bmatrix} \dot{z}_1 \\ \dot{z}_2 \end{bmatrix} = \begin{bmatrix} L_f h_2(x) \\ L_f h_1(x) \end{bmatrix} + \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \hat{u}_{dr} \\ \hat{u}_{qr} \end{bmatrix}. \tag{15}$$

By defining errors

$$\begin{aligned} e_{z1} &= z_1 - \psi_r^{*2} \\ e_{z2} &= z_2 - T_e^*, \end{aligned} \tag{16}$$

derivation of error model (16) gives

$$\begin{aligned} \dot{e}_{z1} &= L_f h_2(x) + \hat{u}_{dr} - 2\psi_r^* \dot{\psi}_r^* \\ \dot{e}_{z2} &= L_f h_1(x) + \hat{u}_{qr} - T_e^*, \end{aligned} \tag{17}$$

and its error dynamics are derived as follows

$$\dot{e}_z = A(x) + B(x)\hat{U} \,, \tag{18}$$

where

$$A(x) = \begin{bmatrix} L_f h_2(x) \\ L_f h_1(x) \end{bmatrix}, \quad B(x) = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \tag{19}$$

considering error model (18) as:

$$\dot{e}_{z1} = L_f h_2(x) + \hat{u}_{dr} + \phi_1 d_1(x)$$
$$\dot{e}_{z2} = L_f h_1(x) + \hat{u}_{qr} + \phi_2 d_2(x)$$
$$\dot{e}_z = \left[A(x) + \Delta A(x) \right] + B(x)\hat{U} \,, \tag{20}$$

where ϕ_i ($i = 1,2$) and $\Delta A(x)$ denote the uncertainties defined

$$\Delta A(x) = \begin{bmatrix} L_{\Delta f} h_2(x) \\ L_{\Delta f} h_1(x) \end{bmatrix} \tag{21}$$

with

$$\Delta f(x) = \begin{bmatrix} -\Delta\alpha i_{ds} + \Delta\beta\psi_{dr} \\ -\Delta\alpha i_{qs} + \Delta\beta\psi_{qr} \\ \dfrac{\Delta R_r L_m}{L_r} i_{ds} - \dfrac{\Delta R_r}{L_r}\psi_{dr} \\ \dfrac{\Delta R_r L_m}{L_r} i_{qs} - \dfrac{\Delta R_r}{L_r}\psi_{qr} \end{bmatrix} \tag{22}$$

and

$$\alpha = \left(\frac{R_s}{L_\sigma} + \frac{R_r L_m{}^2}{L_r{}^2 L_\sigma} \right), \ \beta = \frac{R_r L_m}{L_r{}^2 L_\sigma}, \ \Delta\alpha = \left(\frac{\Delta R_s}{L_\sigma} + \frac{\Delta R_r L_m{}^2}{L_r{}^2 L_\sigma} \right), \ \Delta\beta = \frac{\Delta R_r L_m}{L_r{}^2 L_\sigma}. \tag{23}$$

Therefore

$$\begin{aligned} L_{\Delta f} h_2(x) &= \frac{2\Delta R_r L_m}{L_r}\left[i_{ds}\psi_{dr} + i_{qs}\psi_{qr} \right] - \frac{2\Delta R_r}{L_r}\left[\psi_{dr}^2 + \psi_{qr}^2 \right] \\ L_{\Delta f} h_1(x) &= \frac{3P}{2}\frac{L_m}{L_r}\left(-\frac{\Delta R_r}{L_r}(\psi_{dr} i_{qs} - \psi_{qr} i_{ds}) - \Delta\alpha(i_{qs}\psi_{dr} - i_{ds}\psi_{qr}) \right) \end{aligned} \tag{24}$$

and

$$\begin{aligned} \phi_1 d_1(x) &= \frac{2\Delta R_r}{L_r}\left[L_m(i_{ds}\psi_{dr} + i_{qs}\psi_{qr}) - (\psi_{dr}^2 + \psi_{qr}^2)\right] \\ \phi_2 d_2(x) &= -\frac{3P}{2}\frac{L_m}{L_r}\left[\frac{\Delta R_r}{L_r} + \Delta\alpha\right]\left[\psi_{dr} i_{qs} - \psi_{qr} i_{ds}\right], \end{aligned} \tag{25}$$

that is $\left[\phi_1 d_1(x), \phi_2 d_2(x)\right]^T = \Delta A(x)$.

Since the system resistances R_r and R_s are sensitive to the thermal drift, we assumed that $|\phi_i|$ $(i = 1,2)$ is the unknown and bounded constant. Taking the derivative of (16) with respect to time yields

$$\begin{aligned} \dot{e}_{z1} &= L_f h_2(x) + \hat{u}_{dr} + \hat{\phi}_1 d_1(x) - (\hat{\phi}_1 - \phi_1) d_1(x) \\ \dot{e}_{z2} &= L_f h_1(x) + \hat{u}_{qr} + \hat{\phi}_2 d_2(x) - (\hat{\phi}_2 - \phi_2) d_2(x), \end{aligned} \tag{26}$$

where $\hat{\phi}_i$ $(i = 1,2)$ is the estimate of ϕ_i, k_i $(i = 1,2)$ is a positive constant feedback gain.

It is obvious that the controllers $\hat{u}_{dr}$ and $\hat{u}_{qr}$ are decoupled with respect to two dynamic models, $\left[e_{z1}, e_{z2}\right]$. According to the equations of (26), the adaptive input-output control for the first equation of (26) is designed as

$$\hat{u}_{dr} = -L_f h_2(x) - \hat{\phi}_1 d_1(x) - k_1 e_{z1} \tag{27}$$

where $k_1 > 0$ and the adaptation law of $\hat{\phi}_1$ is given by

$$\dot{\hat{\phi}}_1 = \gamma_1 e_{z1} d_1(x) \tag{28}$$

where $\gamma_1 > 0$ is the adaptation gain. Similarly, the adaptive input-output feedback control for the second dynamic equation of (23) is designed as follows

$$\hat{u}_{qr} = -L_f h_1(x) - \hat{\phi}_2 d_2(x) - k_2 e_{z2} \tag{29}$$

where k_2 is the positive constant feedback gain and

$$\dot{\hat{\phi}}_2 = \gamma_2 e_{z2} d_2(x) \tag{30}$$

where $\gamma_2 > 0$ is the adaptation gain.

Theorem 2: *Using the controller described by* (27)-(30), *the controlled torque and flux amplitude of the* DFIM *is stable and robust to the mismatched uncertainties due to parameter variations*.

Proof: The proof is obtained by choosing the following Lyapunov function

$$V_1 = \frac{1}{2}\left[{e_{z1}}^2 + {e_{z2}}^2 + \frac{1}{\gamma_1}(\hat{\phi}_1 - \phi_1)^2 + \frac{1}{\gamma_2}(\hat{\phi}_2 - \phi_2)^2 \right] \tag{31}$$

Taking the derivative of (31) with respect to time and then substituting (26) into this derivative, we can obtain

$$\begin{aligned} \dot{V}_1 = e_{z1}\left[L_{f(x)}h_2(x) + \hat{u}_{dr} + \hat{\phi}_1 d_1(x) - (\hat{\phi}_1 - \phi_1)d_1(x) \right] + \\ + e_{z2}\left[L_{f(x)}h_1(x) + \hat{u}_{qr} + \hat{\phi}_2 d_2(x) - (\hat{\phi}_2 - \phi_2)d_2(x) \right] + \\ + \frac{1}{\gamma_1}(\hat{\phi}_1 - \phi_1)\dot{\hat{\phi}}_1 + \frac{1}{\gamma_2}(\hat{\phi}_2 - \phi_2)\dot{\hat{\phi}}_2 \end{aligned} \tag{32}$$

Substituting (27), (29), (28) and (30) into equation (32), gives

$$V_1 = -k_1 e_{z1}^2 - k_2 e_{z2}^2 \le 0 \tag{33}$$

Defining the following equation

$$M(t) = k_1 e_{z1}^2 + k_2 e_{z2}^2 \ge 0 \tag{34}$$

Also, defining the following function

$$V_1(t) = V_1(e(0), \hat{\phi}(0)) + \int_0^t \dot{V}_1(\tau)\,\mathrm{d}\tau = V_1(e(0), \hat{\phi}(0)) - \int_0^t M(\tau)\,\mathrm{d}\tau \tag{35}$$

where $e = \left[e_{z1}, e_{z2}\right]^T$ and $\hat{\phi} = \left[\hat{\phi}_1, \hat{\phi}_2\right]^T$. From the definition of the Lyapanouv function $V_1(t) \ge 0$ and the above equation, the following result can be deduced

$$\lim_{t\to\infty} \int_0^t M(\tau)\,\mathrm{d}\tau \le V_1(e(0), \hat{\phi}(0)) < \infty \tag{36}$$

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

$$M(t) \to 0 \text{ as } t \to \infty \tag{37}$$

That is, e_{z1} and e_{z2} will converge to zero as $t \to \infty$. Therefore, the proposed controller is stable and robust, even if parametric uncertainties exist.

4 Active and Reactive Stator Power Control

For both the motoring and generation modes of DFIM operation, regulation of the stator active-reactive power, whose references are P_s^* and Q_s^*, respectively is desirable. Considering a synchronous d and q axis rotating reference frame with the d axes coinciding with the space voltage vector for the main ac supply, reference [11] shows that:

$$i_q^* = \frac{3}{2}\frac{Q_s^*}{U},$$
$$i_d^* = \frac{3}{2}\frac{P_s^*}{U} \tag{38}$$

In addition the rotor flux references can be obtained as:

$$\psi_d^* = \frac{1}{\beta\omega_0}\left(-\frac{R_s}{\sigma} i_q^* - \omega_0 i_d^*\right)$$
$$\psi_q^* = \frac{1}{\beta\omega_0}\left(\frac{R_s}{\sigma} i_d^* - \omega_0 i_q^* - \frac{1}{\sigma}U\right) \tag{39}$$

So the rotor flux reference and torque reference are calculated as below

$$\psi_r^* = \sqrt{(\psi_d^{*2} + \psi_q^{*2})}$$
$$T_e^* = \mu(\psi_d^* i_q^* - \psi_q^* i_d^*) \tag{40}$$

5 Stabilization of Rotor DC-Link Voltage

The objective of this section is to maintain the rotor dc-link voltage constant during drive system operation, which can be achieved using a supply side three phase PWM converter as shown in Fig. 1. With proper control of this converter the rotor dc link voltage can be maintained constant regardless of the magnitude and direction of the rotor power. Vector-control of this inverter, with a reference frame oriented along the stator (or supply) voltage vector position, enables independent control of active and reactive power flowing between the supply and the supply-side converter [11]. For this purpose a capacitor in the dc-link and supply-side converter is used to mitigate variation of the capacitor voltage due to variation of rotor power.

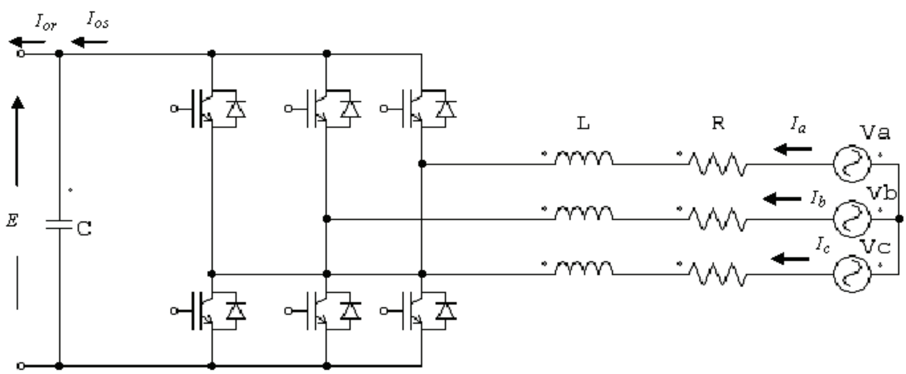

Fig. 1 – *Connection of an inverter to the main supply.*

Considering the d and q axis voltage equation of the main ac supply, derived in the above mentioned rotating reference frame:

$$\begin{aligned}\frac{\mathrm{d}i_d}{\mathrm{d}t} &= \frac{1}{L}(v_d - Ri_d + \omega_e L i_q - v_{d1})\\ \frac{\mathrm{d}i_q}{\mathrm{d}t} &= \frac{1}{L}(-Ri_q - \omega_e L i_d - v_{q1}).\end{aligned} \tag{41}$$

Defining the d and q current errors as

$$\begin{aligned}e_1 &= i_d - i_d^*\\ e_2 &= i_q - i_q^*,\end{aligned} \tag{42}$$

from equations (41) and (42), the d and q current dynamic errors are obtained as

$$\begin{aligned}\dot{e}_1 &= \frac{v_d}{L} - \frac{R}{L}i_d + \omega_e i_q - \frac{v_{d1}}{L} - \dot{i}_d^*\\ \dot{e}_2 &= -\frac{R}{L}i_q - \omega_e i_d - \frac{v_{q1}}{L} - \dot{i}_q^*.\end{aligned} \tag{43}$$

Introducing the following control efforts

$$\begin{aligned}v_{d1} &= L\left(\frac{v_d}{L} - \frac{R}{L}i_d + \omega_e i_q - \dot{i}_d^* - ke_1\right)\\ v_{q1} &= L\left(-\frac{R}{L}i_q - \omega_e i_d - \dot{i}_q^* - ke_2\right),\end{aligned} \tag{44}$$

we have

$$\begin{aligned}\dot{e}_1 &= -ke_1\\ \dot{e}_2 &= -ke_2.\end{aligned} \tag{45}$$

Considering the Lyapanouv candidate as

$$V = \frac{1}{2}e_1^2 + \frac{1}{2}e_2^2 \tag{46}$$

Derivations of function ∇ with respect to time t give

$$\dot{V} = \dot{e}_1 e_1 + \dot{e}_2 e_2 = -k(e_1^2 + e_2^2) < 0 \tag{47}$$

The strategy is shown in Fig. 2.

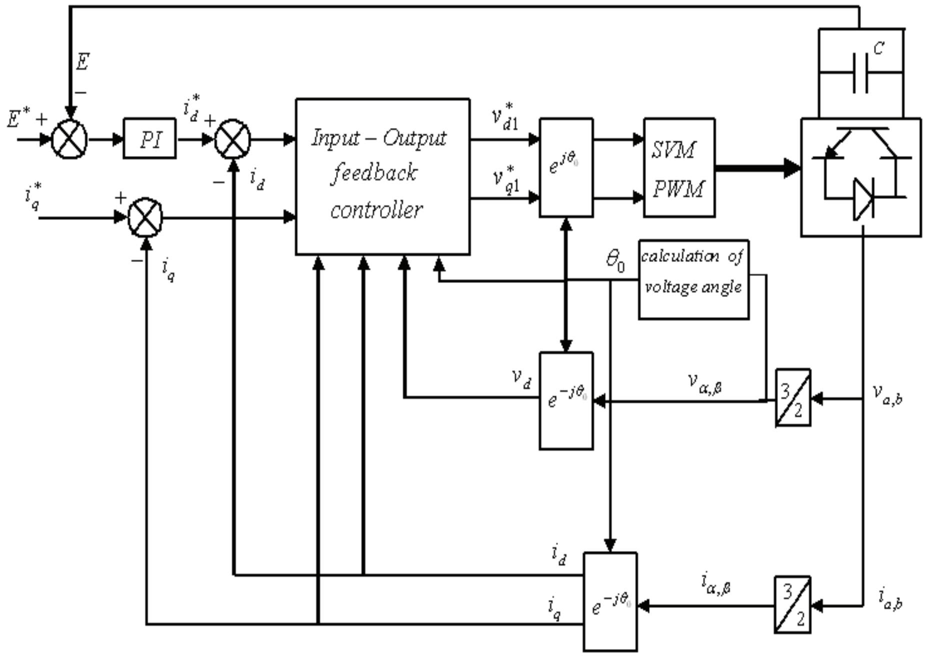

Fig. 2 – DC *link voltage controller.*

I_{or} acting as a disturbance and its effect in the simulation.

Neglecting losses in the supply side converter the injected active power to the rotor is:

$$P_r = Ei_{or} \tag{48}$$

also from Fig. 1, one can obtain

$$\frac{3}{2}v_d i_d = Ei_{os} \tag{49}$$

$$C\frac{\mathrm{d}E}{\mathrm{d}t} = i_{os} - i_{or} \tag{50}$$

Substituting i_{or} and i_{os} from equations (48) and (49) into (50), the rotor dc link voltage variation is obtained as:

$$\mathrm{d}E = \frac{1}{C}\left(\frac{3}{2}\frac{v_d i_d}{E} - \frac{P_r}{E}\right)\mathrm{d}t \tag{51}$$

Based on the theory mentioned in this section, the block diagram shown in Fig. 2 is proposed to maintain the rotor dc link voltage constant.

6 System Simulation

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

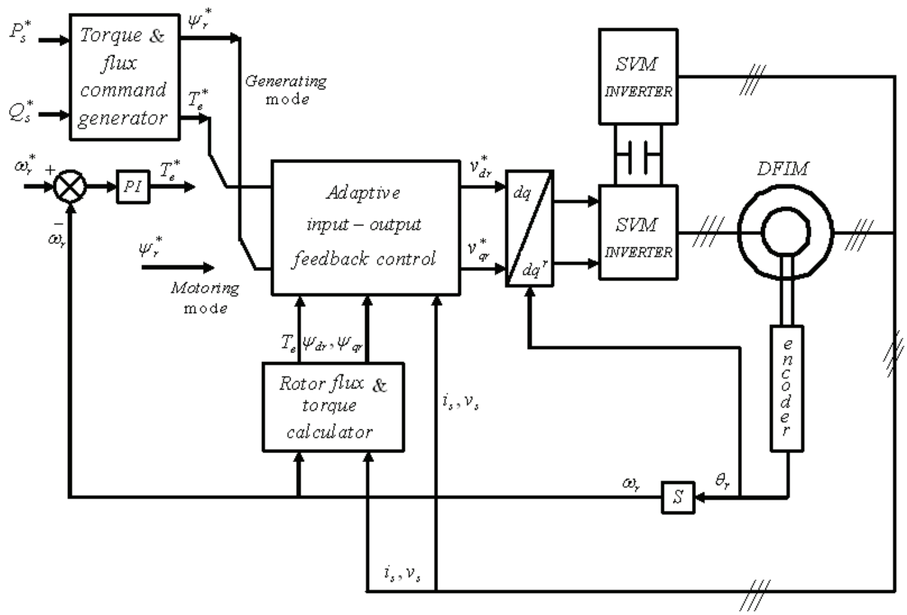

Fig. 3 – *Block diagram of the proposed controller.*

Table 1
Machine Characteristics.

$L_s = 0.094\,\mathrm{H}$	$L_r = 0.022\,\mathrm{H}$	$L_m = 0.041\,\mathrm{H}$
$R_r = 0.45\,\Omega$	$R_s = 0.95\,\Omega$	$J = 0.05\,\mathrm{kg \cdot m^2}$
$T_n = 50\,\mathrm{Nm}$	$2P = 6$	$N_{se} / N_{re} = 2$

In this program, a static Runge-Kutta fourth order method is used to solve the system of 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 by simulation. The machine characteristics are given in **Table 1**.

Simulation results shown in Fig. 4 are obtained for the system motoring mode of operation below and above the synchronous speed. These results are obtained for the condition of an exponential speed reference from 0 to

290 (rad/s) rise up to 375 (rad/s) at $t = 2\text{s}$ with $\tau_\omega = 0.07$, an exponential reference flux signal from zero to 0.3W-t at $t = 0$, down to 0.235W-t at $t = 2\text{s}$ with $\tau_f = 0.01$, a torque reference profile shown in Fig. (4-3) and $R_r = 2R_{rn}$, $R_s = 2R_{sn}$, where R_s and R_r are the stator and rotor resistances, respectively. Note that subscript n shows the nominal parameters.

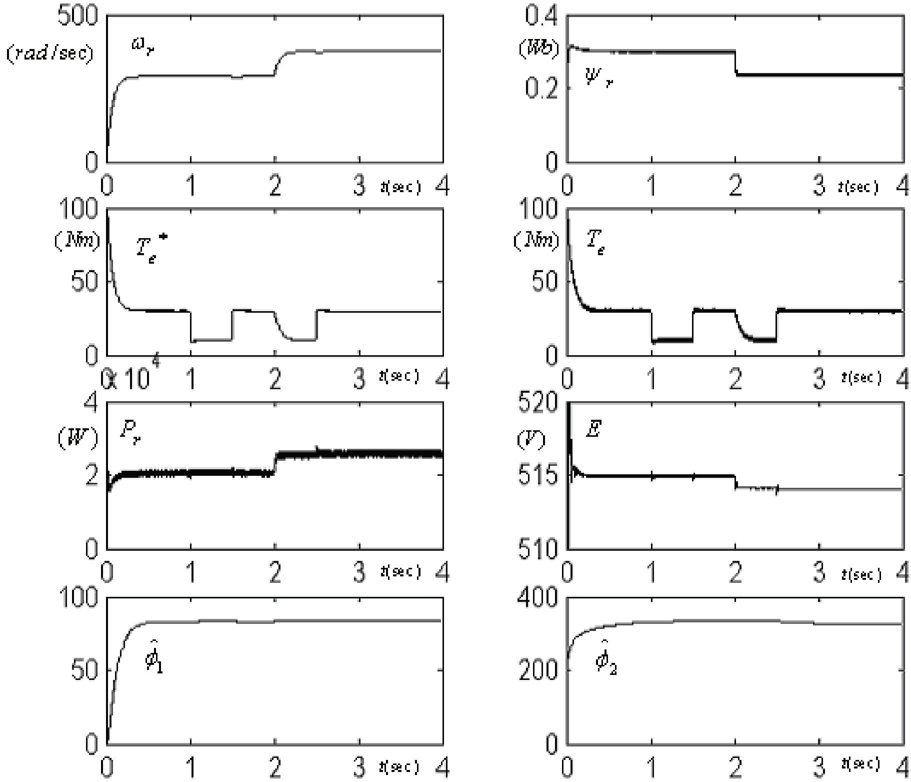

Fig. 4 – *Drive system motoring mode of operation below and above the synchronous speed.*

Fig. 5 shows the drive system performance in the generation mode of operation above the synchronous speed. These results were obtained for the same conditions described for Fig. 4 but for a negative torque reference profile shown in Fig. 5 and a constant speed $\omega_r = 375$ (rad/s).

Fig. 6 shows the drive system performance in the generation mode of operation below the synchronous speed. These results were obtained for the same condition described for Fig. 4 but for a negative torque reference profile shown in Fig. 5 and a constant $\omega_r = 290$ (rad/s).

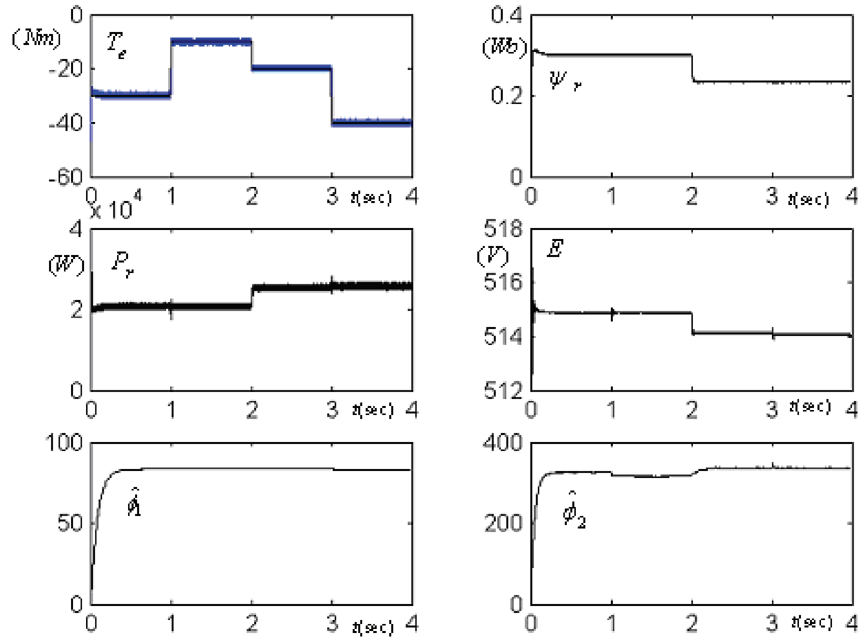

Fig. 5 – *Drive system performance for the generating mode of operation above the synchronous speed.*

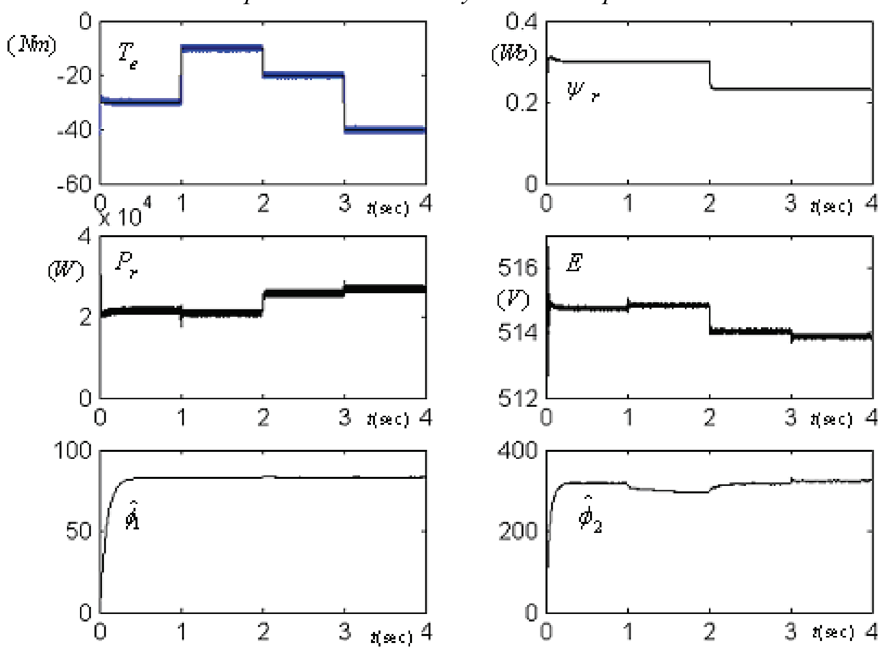

Fig. 6 – *Drive system performance for the generating mode of operation below the synchronous speed.*

7 Conclusion

In this paper an adaptive nonlinear controller has been introduced for DFIM drives. The proposed controller design is based on the adaptive input-output feedback linearization control approach and is capable of making the system states trajectories follow the torque and flux reference signals in spite of stator and rotor resistance uncertainties and external load torque disturbance. The proposed control approach has been tested for both the motoring and generating modes of operation bellow and above the synchronous speed, using two level SVM-PWM back-to-back voltage source inverters in the rotor circuit. Furthermore the rotor dc link voltage is maintained constant also based on the input-output control method, using a rotating synchronous reference frame with d axis coinciding with the direction of the space voltage vector of the main ac supply. Computer simulation results obtained, confirm the validity and effectiveness of the proposed control approach.

8 References

[1] D. Casrdei, F. Profumo, G. Serra, A. Tani: FOC and DTC: Tow Viable Schemes for Induction Motors Torque Control, IEEE Trans. Power Electron., Vol.17, No. 5, Sept. 2002, pp. 779-787.

[2] B. Hopfensperger, J. Atkinson, R.A. Lakin: Stator-flux-oriented Control of a Doubly-fed Induction Machine With and Without Position Encoder, IEE Proc. Electr. Power Appl., Vol. 147, No. 4, July. 2000, pp. 241-250.

[3] L. Xu, W. Cheng: Torque and Reactive Power Control of a Doubly-fed Wound Rotor Induction Machine by Position Sensorless Scheme, IEEE Trans, Ind. Appl., 1995, 31(3), pp. 636-642.

[4] R. Pena, J.C. Clare, G.M. Asher: Doubly Fed Induction Generator Using Back To Back PWM Converter and its Application to Variable-speed Wind Energy Generator, IEE Proc. Electr. Power. Appl., Vol. A3, No. 3, Nov. 1996, pp. 231-241.

[5] Z. Wang, F. Wang, M. Zong, F. Zhang: A New Control Strategy by Combining Direct Torque Control with Vector Control for Doubly Fed Machines, in Proc. IEEE-Powercon' 2004, Singapore, 21-24 Nov. 2004, pp. 792-795.

[6] S. Peresada, A. Tilli, A. Tonielli: Robust Active-reactive Power Control of a Doubly-fed Induction Generator, in Proc. IEEE-IECON '98, Aachen, Germany, Sept. 1998, pp. 1621-1625.

[7] S. Peresada, A. Tilli, A.Tonielli: Indirect Stator Flux-oriented Output Feedback Control of a Doubly Fed Induction Machine, IEEE Trans. Contr. Sys., Vol. 11, No. 6, Nov. 2003, pp. 875 - 888

[8] H.J. Shieh, K.K. Shyu: Nonlinear Sliding-Mode Torque Control with Adaptive Backstepping Approach for Induction Motor Drive, IEEE Trans, Ind. Electr, Vol. 46, No. 2, April 1999, pp. 380-389.

[9] R.D. Lorenz, T.A. Lipo, D.W. Novotny: Motion Control with Induction Motors, Proc. IEEE, Vol. 82, Aug. 1994, pp. 1215-1240.

[10] M. Krstic, I. Kanellakopoulos, P. Kokotovic: Nonlinear and Adaptive Control Design New York, Wiley, 1995.

[11] S. Peresada, A. Tilli, A. Tonielli: Robust Output Feedback Control of a Doubly Fed Induction Machine, Proc. IEEE, 1999, pp. 1348-1354.