

Robust Backstepping Speed Controller of a Doubly Fed Induction Motor using State-Space Nonlinear Approach

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Abstract: - The aim of this paper is an efficiency Backstepping speed controller with direct stator flux orientation of doubly fed induction motor (DFIM) fed by two PWM inverters with separate DC bus link, the Backstepping controller is designed based on the Lyapunov stability theorem, by introducing the approach for decoupling the motor's currents in a rotating (d-q) frame, based on the state space input-output decoupling method. The purpose is therefore to make the speed and the flux control resist to parameter variations, because the variation of parameters during motor operation degrades the performance of the controllers. The proposed approach is analysed by Simulink/Matlab environment. The simulation results show good performance and robustness.

Key-Words: - Doubly fed induction motor (DFIM), backstepping control, state-space nonlinear approach

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1 Introduction

Due to its good performance, robustness and easy maintenance, simplicity of construction, low cost, reliability, the doubly fed induction machine (DFIM) can provide a very attractive solution especially for variable-speed applications such as wind turbine systems, marine propulsion and electric vehicles [1], [2], [3], [4]. The control of the doubly fed induction machine (DFIM) is challenging, since the dynamical system is multivariable, coupled, and highly nonlinear due to the coupling between the flux and the electromagnetic torque [5], [6], [7]. The theory of decoupling by state feedback consists in ensuring an input-output decoupling of a multivariable system by means of a state feedback. This technique is of great theoretical and practical importance insofar as it seeks to break down a multivariable system into several monovariable subsystems with similar dynamics. This decoupling method is of particular interest to DFIM. However, the performance is sensitive to the variations of machine parameters, because the control laws using the PI type controllers give good results in the case of linear systems with constant parameters, but for nonlinear systems, these conventional control laws can be insufficient because they are not robust especially when the requirements on the speed and other dynamic characteristics of the system are strict [8], [9]. In order to improve the performance of the

DFIM control and make it insensitive to parameter variations, and disturbances, we propose the Backstepping speed controller. Backstepping control is an efficient method for nonlinear system. In the backstepping procedure, the first step is to define a virtual control state and then it is forced to become a stabilizing function. Consequently, by appropriately designing the related control input on the basis of Lyapunov stability theory, the error variable can be stabilized. Based on the backstepping design principle, the overall controller design can be established [10], [11].

This paper is organized as follows: principle of field-oriented control by decoupling state space is given in section 2, with his application to the DFIM. The Backstepping speed controller is presented in section 3. In section 4, results of simulation tests are reported. Finally, section 5 draws conclusions.

2 Vector Control Strategy of DFIM by Decoupling State Space

To independently control the electromagnetic force and the flux of DFIM it is necessary to make a judicious choice of reference. For this, one places oneself in a reference frame (d-q) linked to the rotating field with an orientation of the stator flux (the axis d aligned with the direction of the stator flux), as the following figure shows

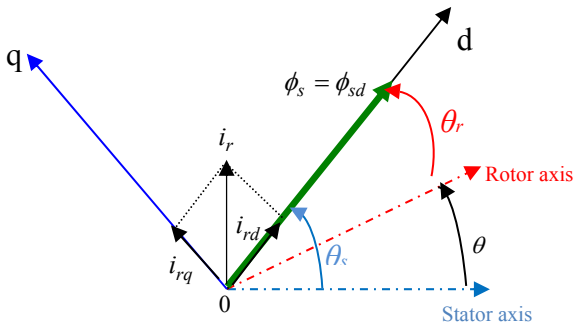


Figure 1: Stator field orientation on the d-axis

We obtain :

$$\phi_{sd} = \phi_s \text{ and } \phi_{sq} = 0$$

By imposing ($i_{sd-ref} = 0$) to ensure a unitary power factor working [12], [13].

Dynamic Model of DFIM:

$$\begin{cases} V_{sd} = R_s \cdot i_{sd} + \frac{d\phi_{sd}}{dt} - \left(\frac{d\theta_s}{dt} \right) \phi_{sq} \\ V_{sq} = R_s \cdot i_{sq} + \frac{d\phi_{sq}}{dt} - \left(\frac{d\theta_s}{dt} \right) \phi_{sd} \\ V_{rd} = R_r \cdot i_{rd} + \frac{d\phi_{rd}}{dt} - \left(\frac{d\theta_r}{dt} \right) \phi_{rq} \\ V_{rq} = R_r \cdot i_{rq} + \frac{d\phi_{rq}}{dt} - \left(\frac{d\theta_r}{dt} \right) \phi_{rd} \end{cases} \quad (1)$$

The fluxes are given by:

$$\begin{cases} \phi_{sd} = L_s i_{sd} + L_m i_{rd} \\ \phi_{sq} = L_s i_{sq} + L_m i_{rq} \\ \phi_{rd} = L_r i_{rd} + L_m i_{sd} \\ \phi_{rq} = L_r i_{rq} + L_m i_{sq} \end{cases} \quad (2)$$

We obtain :

$$\begin{cases} i_{sd} = \frac{1}{L_s} (\phi_{sd} - L_m i_{rd}) \\ i_{sq} = \frac{1}{L_s} (\phi_{sq} - L_m i_{rq}) \end{cases} \quad (3)$$

The electromagnetic torque is expressed by:

$$T_{em} = \frac{pL_m}{L_s} (\phi_{sq} i_{rd} - \phi_{sd} i_{rq}) \quad (4)$$

Replacing equation (1) in (2) and (4) we find:

$$T_{em} = \frac{pL_m}{L_s} (-\phi_s i_{rq}) = -\frac{pL_m}{L_s} \phi_s i_{rq} \quad (5)$$

and

$$\begin{cases} \phi_{sq} = 0 \Rightarrow i_{sq} = -\frac{L_m}{L_s} i_{rq} \\ i_{sd} = 0 \\ i_{rd} = \frac{\phi_s^*}{L_m} \end{cases} \quad (6)$$

$$T_{em} = p L_m i_{rd} i_{sq} \quad (6)$$

2.1 Currents Decoupling by State Space

2.1.1 Principle of the method

Consider the following multivariable system:

$$\begin{cases} \dot{x} = Ax + Bu \\ y = Cx \end{cases} \quad (7)$$

with $x \in \mathfrak{R}^n$, $y \in \mathfrak{R}^m$, $u \in \mathfrak{R}^m$

The objective is to determine a state space of the form:

$$u = -K_d x + L_d v \quad (8)$$

with $v \in \mathfrak{R}^m$

v denotes the new input vector, which decouples the system, in a way that the output y_i (1 to m) depends only on the input v . The output y_i is written:

$$y_i = C_i x \quad (9)$$

where C_i is the i th row of the m matrix C . Let us derive y_i a few times in order to bring up the command. We call characteristic index noted δ_i , the number of derivation it takes in order to bring up the command.

We then have successively for each output i [14], [15]:

$$\begin{cases} \dot{y}_i = C_i \dot{x} = C_i (Ax + Bu) = C_i Ax \\ \ddot{y}_i = C_i A \dot{x} = C_i A (Ax + Bu) = C_i A^2 x \\ \ddot{\ddot{y}}_i^{(3)} = C_i A^2 \dot{x} = C_i A^2 (Ax + Bu) = C_i A^3 x \\ \vdots \\ y_i^{(\delta_i)} = C_i A^{\delta_i} x = C_i A^{\delta_i-1} Bu = C_i Ax \end{cases} \quad (10)$$

Where $C_i A^{\delta_i} B \neq 0$ gives us the value of δ_i

That we can still write in matrix form:

$$\begin{bmatrix} y_1^{(\delta_1)} \\ y_2^{(\delta_2)} \\ \vdots \\ y_m^{(\delta_m)} \end{bmatrix} = \begin{bmatrix} C_1 A^{\delta_1} \\ C_2 A^{\delta_2} \\ \vdots \\ C_m A^{\delta_m} \end{bmatrix} x + \begin{bmatrix} C_1 A^{\delta_1-1} B \\ C_2 A^{\delta_2-1} B \\ \vdots \\ C_m A^{\delta_m-1} B \end{bmatrix} u \quad (11)$$

That is:

$$y' = A'x + B'u \quad (12)$$

with $y \in \mathfrak{R}^m$, $A' \in \mathfrak{R}^{m \times m}$ and $B' \in \mathfrak{R}^{m \times m}$

We seek a control law:

$$u = -K_d x + L_d v \text{ such as } y' = v$$

The looped system is written:

$$\begin{aligned} y' &= A'x + B'(-K_d x + L_d v) \\ &= (A' - B'K_d)x + B'L_d v \end{aligned} \quad (13)$$

To obtain $y' = v$ we must have $B'L_d = 1$ and $A' - B'K_d = 0$

If the matrix B' is invertible, the choice of :

$$K_d = (B')^{-1} A' \text{ and } L_d = (B')^{-1}$$

Gives: $y' = v$

$$\text{That is: } Y_i(s) = \frac{1}{s^{\delta_i+1}} V_i(s)$$

2.2 Application to the DFIM

We search to exploit this method for decoupling the currents of the machine projected on a (d-q) rotating frame [16]. Starting from choosing a state vector equal to the output vector, formed of four currents of the machine. The input vector is formed of supply voltages. Then we obtain the following expression:

$$\begin{cases} \dot{x} = Ax + Bu \\ y = Cx \end{cases} \quad (14)$$

with:

$$x = \begin{bmatrix} I_{sd} & I_{sq} & I_{rd} & I_{rq} \end{bmatrix}^T \text{ the state vector}$$

$$u = \begin{bmatrix} V_{sd} & V_{sq} & V_{rd} & V_{rq} \end{bmatrix}^T \text{ the input vector}$$

voltages

$$A = \begin{bmatrix} -a_1 & \omega + \omega_s & a_3 & a_5 \omega \\ -a\omega + \omega_s & -a_1 & -a_5 \omega & a_3 \\ a_4 & -a_6 \omega & -a_2 & -\frac{\omega}{\sigma} + \omega_s \\ a_6 \omega & a_4 & \frac{\omega}{\sigma} - \omega_s & -a_2 \end{bmatrix}$$

$$B = \begin{bmatrix} b_1 & 0 & -b_3 & 0 \\ 0 & b_1 & 0 & -b_3 \\ -b_3 & 0 & b_2 & 0 \\ 0 & -b_3 & 0 & b_2 \end{bmatrix}$$

$$C = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} = I_4$$

$$a = \frac{1-\sigma}{\sigma}; a_1 = \frac{R_s}{\sigma L_s}; a_2 = \frac{R_r}{\sigma L_r}; a_3 = \frac{R_r L_m}{\sigma L_s L_r};$$

$$a_4 = \frac{R_s L_m}{\sigma L_s L_r}; a_5 = \frac{L_m}{\sigma L_s}; a_6 = \frac{L_m}{\sigma L_r}; b_1 = \frac{1}{\sigma L_s};$$

$$b_2 = \frac{1}{\sigma L_r}; b_3 = \frac{L_m}{\sigma L_s L_r}; \sigma = 1 - \frac{L_m^2}{L_s L_r}$$

The choice of $x = y$ makes the system completely controllable and observable. In applying the decoupling method on this system, it follows that:

$$\forall i; \delta_i = 0 \text{ and } \begin{cases} L_d = B^{-1} \\ K_d = B^{-1} A \end{cases}$$

$$\text{where: } L_d = \begin{bmatrix} L_s & 0 & L_m & 0 \\ 0 & L_s & 0 & L_m \\ v & 0 & L_r & 0 \\ 0 & L_m & 0 & L_r \end{bmatrix}$$

and:

$$K_d = \begin{bmatrix} -R_s & L_s \omega & 0 & L_m \omega_s \\ -L_s \omega_s & -R_s & -L_m \omega_s & 0 \\ 0 & M_{sr}(\omega_s - \omega) & -R_r & L_r(\omega_s - \omega) \\ -L_m(\omega_s - \omega) & 0 & -L_r(\omega_s - \omega) & -R_r \end{bmatrix}$$

$$\dot{y} = u^*, \text{ therefore: } \frac{Yi(s)}{u^*(s)} = \frac{1}{s}$$

The four currents are decoupled and thus governed by the same transfer function in open loop

$$G(s) = \frac{1}{s}$$

2.2.1 Design the Currents Control Loops

The currents are decoupled, and then we can consider a state space correction with the method of placement of poles. The principal schematic diagram of this correction is given by the figure 2.

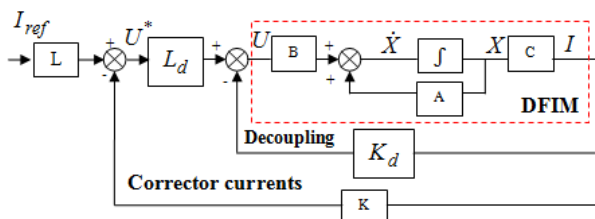


Figure 2: Current regulation by state space

To ensure the same response for the current loop, the next choice can be adopted.

$$L = K = \begin{bmatrix} K & 0 & 0 & 0 \\ 0 & K & 0 & 0 \\ 0 & 0 & K & 0 \\ 0 & 0 & 0 & K \end{bmatrix} \quad (15)$$

So the transfer function of each current closed loop will be of the form:

$$H(s) = \frac{K}{K + S} \quad (16)$$

3 Backstepping Speed Controller

The nonlinear systems control domain requires high performance algorithms to ensure a better level of stability and performance. The Backstepping control is part of the algorithms in this area. This technique, which was developed by Kanellakopoulos in 1990, offers a systematic method for producing controllers for nonlinear systems [17], [18], [19]. In this section, a robust backstepping controller design method based on the Lyapunov stability technique is developed for speed control of a doubly fed induction motor. The main control objective is to force the motor speed to asymptotically track a desired reference speed while simultaneously rejecting the external disturbances. Consequently, this control thus helping it possible to operate the complete system in the best performances in the static and dynamic regimes.

The proposed design procedure includes the steps which are elaborately discussed as follows.

From this equation, it is not difficult to drive:

$$\frac{d\Omega(t)}{dt} = \frac{1}{J} [T_{em}(t) + T_L(t) + f\Omega(t)] \quad (17)$$

where:

$$\frac{d\Omega(t)}{dt} = aT_{em}(t) + bT_L(t) + c\Omega(t) \quad (18)$$

With: a, b and c are constant parameters which are related to the motor parameters.

$$a = \frac{1}{J}, \quad b = -\frac{1}{J}, \quad c = \frac{f}{J}$$

The first step of the Backstepping control is defined the speed track error:

$$e(t) = \Omega_{ref}(t) - \Omega(t) \quad (19)$$

Then the derivative of speed track error can be represented as:

$$\dot{e}(t) = \dot{\Omega}_{ref}(t) - \dot{\Omega}(t) \quad (20)$$

with:

$$\dot{\Omega}(t) = aT_{em}(t) + bT_L(t) + c\Omega(t) \quad (21)$$

Then:

$$\dot{e} = \dot{\Omega}_{ref}(t) - aT_{em}(t) - bT_L(t) - c\Omega(t) \quad (22)$$

Subsequently we define the Lyapunov function of the form:

$$V(t) = \frac{1}{2} e^2(t) \quad (23)$$

Its derivative gives:

$$\dot{V}(t) = e(t)\dot{e}(t) = e(t)[\dot{\Omega}_{ref}(t) - aT_{em}(t) - bT_L(t) - c\Omega(t)] \quad (24)$$

In order to ensure the stability of the system, it is necessary to make the derivative of the Lyapunov function

\dot{V} negative. For this, we defined a positive constant 'K' at derivative of equation

$$\dot{V}(t) = -Ke^2(t) \leq 0 \quad (25)$$

The expression of the electromagnetic torque reference T_{em}^* is extracted from equation (25):

$$T_{em}^*(t) = \frac{1}{a} [\dot{\Omega}_{ref}(t) - b.T_L - c.\Omega(t) + K.e(t)] \quad (26)$$

The block diagram of the proposed Backstepping speed controller system is shown in figure 3.

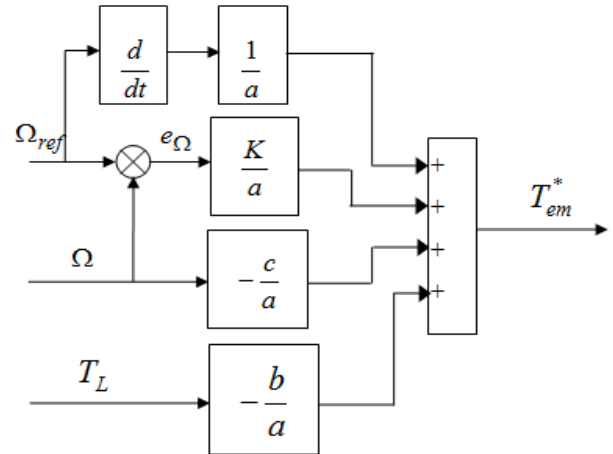


Figure 3: Block diagram of Backstepping speed controller

4 Results and Analysis

To evaluate the performance of the proposed method, the model for the control law applied to the DFIM (Figure 4) has been implemented in the Matlab / Simulink software.

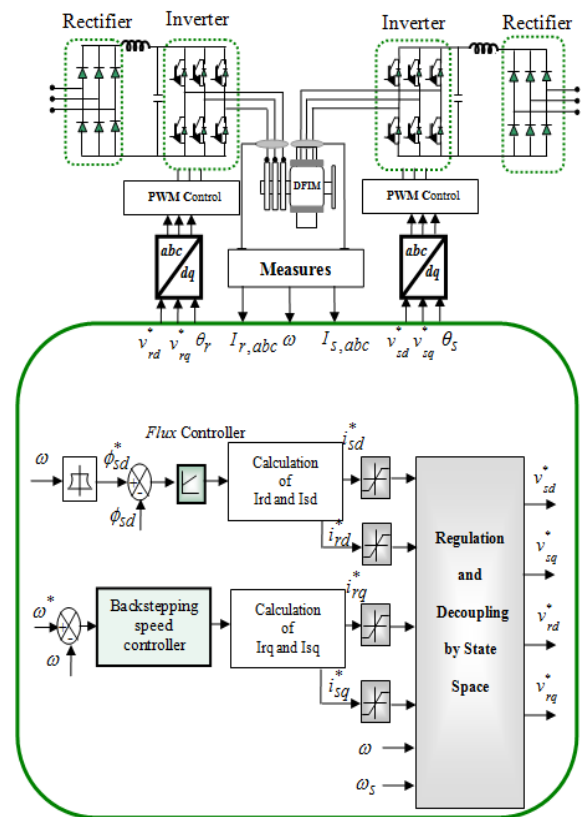


Figure 4: Block Diagram of robust Backstepping speed Controller of DFIM using a state space nonlinear approach

Fig. 4 shows the different steps to turn out the control law of stator and rotor voltage. Accurate knowledge of the rotor speed and rotor flux are the keys factors in obtaining a high-performance and high-efficiency DFIM.

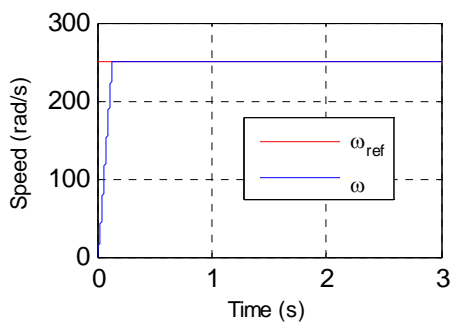
The system parameters of the DFIM tested in this study are given in Appendix.

A. Constant Speed and Load Torque Application

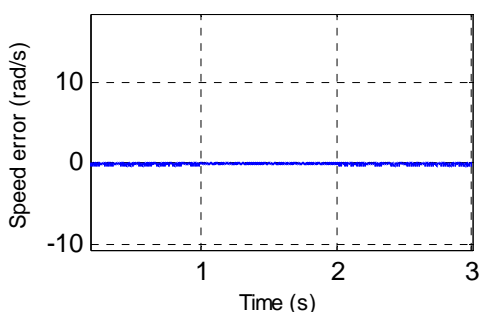
The first test (Figure. 5) concerns a no-load starting of the motor with a reference speed $\omega_{ref} = 250$ rad/s. and a nominal load disturbance torque (10N.m) is suddenly applied between 1sec and 2sec.

From these results, it can be seen that the command using State-Space nonlinear approach has a better regulation (accuracy and stability) of the speed and even the stator flux, since the introduction of the charges has no influence on the evolution (stability) speed and also the flow, which shows the robustness of backstepping speed controller facing these disturbances compared to the conventional PI regulator.

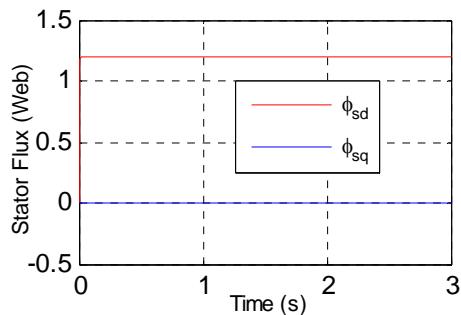
This simulation results clearly show that, the decoupling between the electromagnetic torque and the stator flux is very satisfactory.



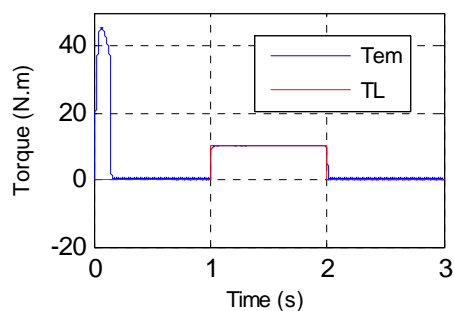
(a)



(b)



(c)



(d)

Figure 5: Simulation waveforms of proposed nonlinear control drives: **a.** Reference, Mesure Rotor speed, **b.** Zoom of Speed Error, **c.** Stator Flux, **d.** Electromagnetic torque waveform.

B. Robustness test

1. Inverse Rotation Speed

Figure 6 shows the evolution of the characteristics of the DFIM with speed control by backstepping controller, rotor and stator currents based on the input-output decoupling method, followed by the inversion of the speed from 250 to -250 rad / s, with a load torque of 10N.m is suddenly applied between 1sec and 2sec. As it's shown by figure a.6, the rotor speed tracks perfectly their reference with a static error equal to zero (Figure. 6. b).

This results express that the effect produced by the load torque variation is almost negligible for the system. also we notice decoupling between the electromagnetic torque and the stator flux is very satisfactory

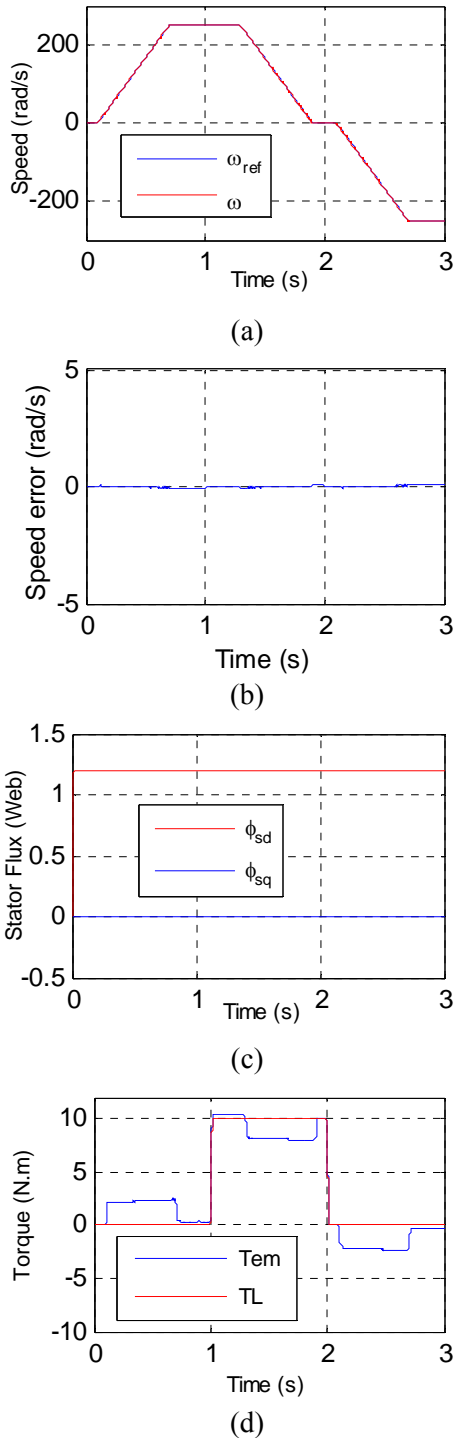


Figure 6: Simulation results of proposed nonlinear control drives with inverse rotation speed: a. Reference, Measure Rotor speed, b. Speed Error, c. Stator Flux, d. Electromagnetic torque waveform.

2. The rotor resistance variation test

In order to study the influence of parametric variations on the behavior of the nonlinear control, we introduced a variation of + 50% of R_r in the first test, we obtained the results as shown in the

following figures:

According to the result of Figure. 7, we note that the increase in resistance did not affect the accuracy of a backstepping speed controller, we clearly see that the rotor speed perfectly follows its reference. An increase of the rotor resistance gives best performances.

The results of the speed control have shown that the control with backstepping speed controller ensures good performance even in the presence of parametric variations and external disturbances (load disturbance torque). As we see in this results, the increase in resistance did not affect the accuracy and orientation of the stator flux, which proves the robustness of the proposed controller. According to these results, we can say and in general, we obtained the same performance as the previous test with the nominal rotor resistance.

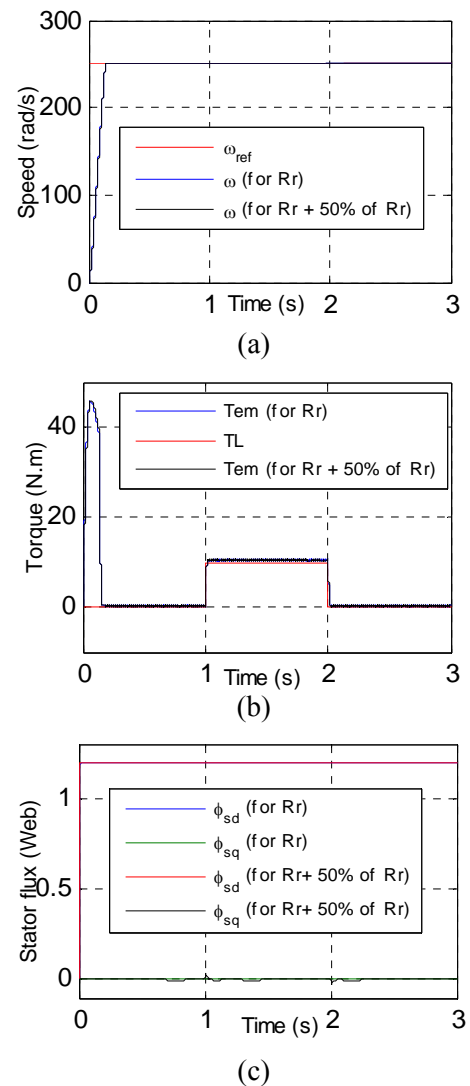


Figure 7: Simulation results of proposed nonlinear control drives under a load change and with rotor resistance increased sharply by 50% from rated value

3. The stator resistance variation test

For a nominal value of R_r , the stator resistance R_s is increased by + 50% of its nominal value, we obtained the results as shown in the following figures: The obtained results (Figure.8) demonstrates that even if the stator resistance changes. This results shows also that using backstepping speed controller performs a better control in terms of robustness

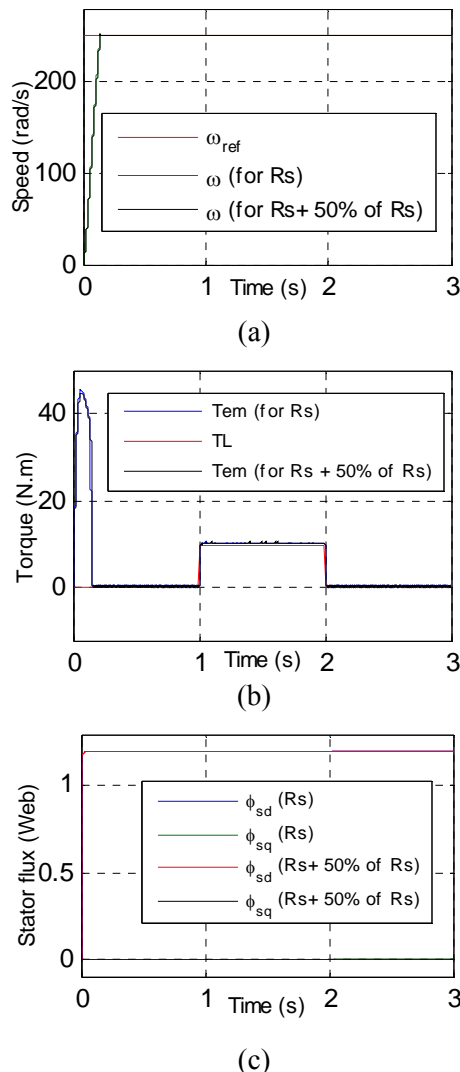


Figure 8: Simulation results of proposed nonlinear control drives under a load change and with stator resistance increased sharply by 50% from rated value

5 Conclusion

This paper deals with decoupling method of the currents for a vector control of a DFIM with stator flux orientation. This approach is based on the state space input-output decoupling, so

we can obtain better decoupling for the currents between d and q axis. the backstepping speed controller was employed to solve different drawbacks of the conventional PI controllers and to obtain the better dynamic performance from the DFIM in a speed control also make it insensitive to parameter variations and disturbances.

Finally, we believe that the proposed solutions will improve the tracking performance of the trajectory and disturbance rejection load Torque and parameter variations and also enhance stability through the robust look of the Backstepping speed controller.

Appendix

DFIM motor parameters

Item	Symbol	Data
DFIM Power	P_w	1.5 Kw
Nominal speed	ω	1450 rpm
Pole pairs number	P	2
Stator resistance	R_s	1.68Ω
Rotor resistance	R_r	1.75Ω
Stator self inductance	L_s	295 mH
Rotor self inductance	L_r	104 mH
Mutual inductance	L_m	165 mH
Moment of inertia	J	0.01 kg.m ²
friction coefficient	F	0.0027kg.m ² /s
Nominal Frequency	f	50 Hz

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