

A High Gain Observer Based Sensorless Nonlinear Control of Induction Machine

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ABSTRACT

In this paper a sensorless Backstepping control scheme for rotor speed and flux control of induction motor drive is proposed. The most interesting feature of this technique is to deal with non-linearity of high-order system by using a virtual control variable to render the system simple. In this technique, the control outputs can be derived step by step through appropriate Lyapunov functions. A high gain observer is performed to estimate non available rotor speed and flux measurements to design the full control scheme of the considered induction motor drive. Simulation results are presented to validate the effectiveness of the proposed sensorless Backstepping control of the considered induction motor.

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1. INTRODUCTION

Induction motor (IM) compared to other types of electric machines, is used in a wide range of industrial applications. This is due to its excellent reliability, great robustness and less maintenance requirements. However, the induction motor model is complicated for various reasons, among them:

- The dynamic behavior of the motor is described by a fifth-order highly coupled and nonlinear differential equations,
- Rotor electric variables (fluxes and currents) are practically unmeasurable state variables,
- Some physical parameters are time-varying (stator and mainly rotor resistance, due to heating, magnetizing induction due to saturation).

The first control schemes of induction motors were based on traditional scalar control that can guarantee only modest performance. In many applications, it is necessary to use more sophisticated controls as Field Oriented Control (FOC) proposed by Blaschke [1]. This type of control technique has led to a radical change in control of the induction machines. Thanks to the quality of dynamic performance that it brings. In the FOC technique, called also vector control, the torque and flux are decoupled by a suitable decoupling network. In this type of control technique, the flux and the torque components are controlled independently by the stator direct and quadratic currents respectively. Thus permits to control the induction motor (IM) as a separately excited DC motor [2]. The high performance of such strategy may be deteriorated in practice due to plant uncertainties.

Other techniques were conceived like input-output linearization technique that is based on the use of differential geometry theory to allow by a diffeomorphic transformation a state feedback control of the induction motor system [3]-[5]. This method cancels the nonlinear terms in the plant model and fails when

the physical parameters are time-varying.

By contrast, the passivity based control doesn't cancel all the nonlinearity terms but ensure system stability, by adding a damping term to the total energy of the system. It is characterized by its robustness against the parameter uncertainties, however, its experimental implementation is still difficult [6], [7]. The sliding mode control is another control technique that is characterized by simplicity of design and attractive robustness properties. Its major drawback is the chattering phenomenon [8]-[10].

Since last two decades, the nonlinear control called "Backstepping" became one of the most popular control techniques for a wide range of nonlinear system classes [13]-[21]. It is distinguished by its ability to easily guarantee the global stabilization of system, even in the presence of parametric uncertainties [18]. The design of the control law is based mainly on the construction of appropriate Lyapunov functions. Its present form is due to Krstic, Anellakopoulos and Kokotovic [13] based on the Lyapunov stability tools, this approach offers great flexibility in the synthesis of the regulator and naturally leads itself to an adaptive extension case. This control technique offers good performance in both steady state and transient operations, even in the presence of parameter variations and load torque disturbances.

In order to implement a nonlinear sensorless control technique, to improve the robustness and the reliability of induction motor drives, it is necessary to synthesize a state observer for the estimation of non-measurable state variables of the machine system that are essential for control purposes [23]-[26]. Among the observation techniques one can use the high gain observer technique to design an appropriate sensorless control of IM drives.

In this paper a Backstepping control that involves non measurable state variables of the induction motor system is performed. In order to achieve a sensorless control a high gain observer is designed to estimate non measured state variable of the machine.

The paper is organized as follows: In section two the nonlinear induction motor model is presented. Backstepping speed and flux controllers design is presented in section three. The high gain observer technique is presented in the section four. In the fifth and final section simulation results and comment are presented.

2. INDUCTION MOTOR NONLINEAR MODEL

In order to reduce the complexity of the three phase induction motor model, an equivalent two phase representation is used under assumptions of linearity of the magnetic circuit and neglecting iron losses. This type of model is designed in the fixed stator reference frame (α, β).

In this paper, the considered induction motor model has stator current, rotor flux and rotor angular velocity as selected state variables. The control inputs are the stator voltage and load torque. The available induction motor stator current measurements are retained as the motor system outputs.

In these conditions, the nonlinear model of the induction motor can be expressed as the following"

$$\dot{x} = f(x) + g(x)u \quad (1)$$

$$y = h(x) \quad (2)$$

With,

$$f(x) = \begin{bmatrix} -\gamma i_{s\alpha} + \frac{K}{T_r} \varphi_{r\alpha} + p\omega \varphi_{r\beta} \\ -\gamma i_{s\beta} - p\omega K \varphi_{r\alpha} + \frac{K}{T_r} \varphi_{r\beta} \\ \frac{M}{T_r} i_{s\alpha} - \frac{1}{T_r} \varphi_{r\alpha} + p\omega \varphi_{r\beta} \\ \frac{M}{T_r} i_{s\beta} - p\omega \varphi_{r\alpha} - \frac{1}{T_r} \varphi_{r\beta} \\ \frac{pM}{jL_r} (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) - \frac{f}{j} \Omega - \frac{T_l}{j} \end{bmatrix} \quad (3)$$

$$g(x) = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 & 0 & 0 \\ 0 & \frac{1}{\sigma L_s} & 0 & 0 \end{bmatrix}^T \quad (4)$$

$$\text{And } h(x) = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \quad \text{With: } \gamma = \frac{R_s}{\sigma L_s} + \frac{R_r M^2}{\sigma L_s L_r^2}, \sigma = 1 - \frac{M^2}{\sigma L_s L_r}, T_r = \frac{L_r}{R_r} \text{ and } K = \frac{M}{\sigma L_s L_r}.$$

Where $x = [i_{s\alpha}, i_{s\beta}, \varphi_{r\alpha}, \varphi_{r\beta}, \omega]^T$ is the state vector and $u = [u_{s\alpha}, u_{s\beta}]^T$ the input vector control; with $i_{s\alpha}, i_{s\beta}$ as the stator currents and $\varphi_{r\alpha}, \varphi_{r\beta}$ as the rotor flux, $u_{s\alpha}, u_{s\beta}$ are the stator command voltages. Ω, R_r, L_r, R_s and L_s are the rotor angular velocity, the rotor resistance, the rotor inductance, the stator resistance and the stator inductance respectively. M is the mutual inductance between stator and rotor winding, p is the number of pair poles, J is the moment of inertia of the rotor, f is the viscous friction coefficient and T_l is the external load torque.

3. SPEED AND FLUX BACKSTEPPING CONTROLLER DESIGN

The Backstepping control design is based on the use of the so-called ‘‘virtual control’’ to systematically decompose a complex nonlinear control problem into simpler one, smaller ones, by dividing the control design into various design steps. In each step we deal with an easier, single-input single-output design problem, and each step provides a reference for the next design step. This approach is different from the conventional feedback linearization in that it can avoid cancellation of useful nonlinearities to achieve the stabilization and tracking objectives.

3.1. First Step

In the first step, it is necessary to specify the desired (reference) trajectories that the system must track, and design controllers to ensure good tracking error.

To this end, we define a reference trajectory, $y_{ref} = (\Omega_{ref}, \varphi_{ref}^2)$, where Ω_{ref} and φ_{ref}^2 are speed and rotor flux modul reference trajectories.

The speed tracking error $e_{1\Omega}$ and the flux magnitude tracking error $e_{1\varphi}$ are defined as:

$$e_{1\Omega} = \Omega_{ref} - \Omega \quad (5)$$

$$e_{1\varphi} = \varphi_{ref}^2 - \varphi_r^2 \quad (6)$$

With $\varphi_r^2 = \varphi_{r\alpha}^2 + \varphi_{r\beta}^2$

The error dynamical equations are:

$$\dot{e}_{1\Omega} = \dot{\Omega}_{ref} - \left[\frac{pM}{jL_r} (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) - \frac{T_l}{j} - \frac{f}{j} \Omega \right] \quad (7)$$

$$\dot{e}_{1\varphi} = \frac{d}{dt} (\varphi_{ref}^2) - \left[\frac{2M}{T_r} (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) \right] + \frac{2}{T_r} \varphi_r^2 \quad (8)$$

By setting the virtual control expressions below:

$$\alpha_1 = \left[\frac{pM}{jL_r} (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) \right] \quad (9)$$

$$\beta_1 = \left[\frac{2M}{T_r} (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) \right] \quad (10)$$

We can write (8) and (9) under the following form:

$$\dot{e}_{1\Omega} = \dot{\Omega}_{ref} - \alpha_1 + \frac{T_l}{j} + \frac{f}{j} \Omega \quad (11)$$

$$\dot{e}_{1\varphi} = \frac{d}{dt} (\varphi_{ref}^2) - \beta_1 + \frac{2}{T_r} \varphi_r^2 \quad (12)$$

Let us check the tracking error dynamics stability by choosing the following candidate Lyapunov function:

$$v_1 = \frac{1}{2} [e_{1\Omega}^2 + e_{1\varphi}^2] \quad (13)$$

The time derivative of (13) gives:

$$\dot{v}_1 = e_{1\Omega} \dot{e}_{1\Omega} + e_{1\varphi} \dot{e}_{1\varphi} \quad (14)$$

To render the time derivative of the Lyapunov function negative definite one has to choose the derivatives of the error tracking as follows:

$$\dot{e}_{1\Omega} = -c_1 e_{1\Omega} \quad (15)$$

$$\dot{e}_{1\varphi} = -d_1 e_{1\varphi} \quad (16)$$

In these conditions the virtual control, deduced from relations (11) and (12) become as the following:

$$\alpha_1 = c_1 e_{1\Omega} + \dot{\Omega}_{ref} + \frac{T_l}{j} + \frac{f}{j} \Omega \quad (17)$$

$$\beta_1 = d_1 e_{1\varphi} + \frac{d}{dt}(\varphi_{ref}^2) + \frac{2}{T_r}(\varphi_{ref}^2 - e_{1\varphi}) \quad (18)$$

Where c_1 and d_1 are the positive design gains that determine the dynamic of closed loop.

The time derivative of the candidate Lyapunov function is evidently negative definite, so the tracking error $e_{1\Omega}$ and $e_{1\varphi}$ can be stabilized.

3.2. Second Step

Previous references, chosen to ensure a stable dynamic of speed and flux tracking error, can't be imposed to the virtual controls without considering errors between them.

To this end, let us define the following errors:

$$e_{2\Omega} = \alpha_1 - \left[\frac{pM}{jL_r} (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) \right] \quad (19)$$

$$e_{2\varphi} = \beta_1 - \left[\frac{2M}{T_r} (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) \right] \quad (20)$$

One determines the new dynamics of the errors $e_{1\Omega}$ and $e_{1\varphi}$, expressed now in terms of $e_{2\Omega}$ and $e_{2\varphi}$.

$$\dot{e}_{1\Omega} = -c_1 e_{1\Omega} + e_{2\Omega} \quad (21)$$

$$\dot{e}_{1\varphi} = -d_1 e_{1\varphi} + e_{2\varphi} \quad (22)$$

From (19) and (20) we obtain the following errors dynamics equations:

$$\dot{e}_{2\Omega} = \alpha_2 + \left[\frac{pK}{j} (\varphi_{r\alpha} u_{s\beta} - \varphi_{r\beta} u_{s\alpha}) \right] \quad (23)$$

$$\dot{e}_{2\varphi} = \beta_2 - \left[2KR_r (\varphi_{r\alpha} u_{s\alpha} + \varphi_{r\beta} u_{s\beta}) \right] \quad (24)$$

Where

$$\alpha_2 = \dot{\alpha}_1 + \frac{pM}{jL_r} \left[\left(\gamma + \frac{1}{T_r} \right) (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) \right] + \frac{pM}{jL_r} \left[p\Omega (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) + K\varphi_r^2 \right]$$

$$\beta_2 = \dot{\beta}_1 + \frac{2M}{T_r} \left[\left(\gamma + \frac{1}{T_r} \right) (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) - \frac{K}{T_r} \varphi_r^2 \right] - \frac{2M}{T_r} \left[p\Omega (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) + \frac{M}{T_r} (i_{s\alpha}^2 + i_{s\beta}^2) \right]$$

One can see, from relations (23) and (24) that the real control components have appeared in the error dynamics. Thus permits us to construct the final Lyapunov function as:

$$v_2 = \frac{1}{2} [e_{1\Omega}^2 + e_{1\varphi}^2 + e_{2\Omega}^2 + e_{2\varphi}^2] \quad (25)$$

So the CLF derivative is determined below, by using (21), (22), (23) and (24):

$$\dot{v}_2 = -c_1 e_{1\Omega}^2 + e_{1\Omega} e_{2\Omega} - d_1 e_{1\varphi}^2 + e_{1\varphi} e_{2\varphi} - c_2 e_{2\Omega}^2 - d_2 e_{2\varphi}^2 + e_{2\Omega} \left(c_2 e_{2\Omega} + \alpha_2 + \frac{pK}{j} (\varphi_{r\alpha} u_{s\beta} - \varphi_{r\beta} u_{s\alpha}) \right) + e_{2\varphi} \left(d_2 e_{2\varphi} + \beta_2 - 2KR_r (\varphi_{r\alpha} u_{s\alpha} + \varphi_{r\beta} u_{s\beta}) \right) \quad (26)$$

Where c_2 and d_2 are the positive design gains that determine the dynamic of closed loop.

In order to make the CLF derivative negative definite as:

$$\dot{v}_2 = -c_1 e_{1\Omega}^2 - d_1 e_{1\varphi}^2 - c_2 e_{2\Omega}^2 - d_2 e_{2\varphi}^2 \leq 0 \quad (27)$$

We choose voltage control as follows:

$$c_2 e_{2\Omega} + \alpha_2 + \frac{pK}{j} (\varphi_{r\alpha} u_{s\beta} - \varphi_{r\beta} u_{s\alpha}) = 0 \quad (28)$$

$$d_1 e_{1\varphi} + \beta_2 - 2KR_r [2KR_r (\varphi_{r\alpha} u_{s\alpha} + \varphi_{r\beta} u_{s\beta})] = 0 \quad (29)$$

Thus leads to the following control expressions:

$$u_{s\alpha} = \frac{1}{\varphi_r^2} \left[\frac{(\beta_2 + e_{2\varphi} + d_2 e_{2\varphi})}{2KR_r} \varphi_{r\alpha} - \frac{j}{pK} [\alpha_2 + e_{1\varphi} + c_2 e_{2\varphi}] \varphi_{r\beta} \right] \quad (30)$$

$$u_{s\beta} = \frac{1}{\varphi_r^2} \left[\frac{(\beta_2 + e_{2\varphi} + d_2 e_{2\varphi})}{2KR_r} \varphi_{r\beta} + \frac{j}{pK} [\alpha_2 + e_{1\varphi} + c_2 e_{2\varphi}] \varphi_{r\alpha} \right] \quad (31)$$

4. NONLINEAR HIGH GAIN OBSERVER DESIGN

Generally, the dynamic behavior of the induction motor (IM) belongs to a class of relatively fast systems. For computational issue, the high gain observer which admits an explicit correction gain can be considered as one of the most viable candidate in the problem of state estimation. Later on, we adopt this method in our design.

Consider the nonlinear uniformly observable class of systems as the following form [23].

$$\dot{x} = f(x, u) + \varepsilon \quad (32)$$

$$y = Cx = x^1 \quad (33)$$

Where the state $x \in \mathbb{R}^n$ with $x^k \in \mathbb{R}^{n_k}$ for $k = 1, 2, \dots, q$ and $n_1 \geq n_2 \geq \dots \geq n_q$. The input $u \in \mathbb{U}$ a compact set of \mathbb{R}^m , the output $y \in \mathbb{R}^{n_1}$.

$$x = \begin{bmatrix} x^1 \\ x^2 \\ \vdots \\ x^q \end{bmatrix}; f(x, u) = \begin{bmatrix} f_1(x^1, x^2, u) \\ f_2(x^1, x^2, x^3, u) \\ \vdots \\ f_{q-1}(x^1, x^2, \dots, x^{q-1}, u) \\ f_q(x, u) \end{bmatrix}; \varepsilon = \begin{bmatrix} 0 \\ \vdots \\ 0 \\ \varepsilon^{q-1} \\ \varepsilon^q \end{bmatrix}$$

With I_{n_1} is the $n_1 \times n_1$ identity matrix and $0_{n_1 \times n_l}$ is the $n_1 \times n_l$ null matrix, $l \in \{2, \dots, q-1\}$, $\varepsilon^k \in \mathbb{R}^{n_k}$, $k \in \{q-1, q\}$, each ε^k is an unknown bounded real valued function that depend on uncertain parameters, in our case we propose $\varepsilon^k = 0$.

The synthesis of the high gain observer (HGO) corresponding to systems of the form (32) and (33), requires making some assumptions as follows:

a) There exist γ, δ with $0 < \gamma \leq \delta$ such that for all $k \in \{1, \dots, q-1\}$, $x \in \mathbb{R}^n$, $u \in U$ we have:

$$0 < \gamma^2 I_{n_k} \leq \left[\frac{\partial f_k(x^{1:k}, u)}{\partial x^{k+1}} \right]^T \frac{\partial f_k(x^{1:k}, u)}{\partial x^{k+1}} \leq \delta^2 I_{n_k}$$

Moreover, we assume that $\text{Rank} \left(\frac{\partial f_k(x^{1:k}, u)}{\partial x^{k+1}} \right) = n_{k+1}$

b) The function $f(x, u)$ is globally Lipschitz with respect to x , uniformly in u .

In these conditions the high gain observer corresponding to systems of the form (32) and (33) can be written as:

$$\dot{\hat{x}} = f(\hat{x}, u) - \theta \Lambda^{-1}(\hat{x}) \Delta_{\theta}^{-1} S^{-1} C^T \bar{C}(\hat{x} - x) \quad (34)$$

Where $\Lambda^{-1}(\hat{x})$ is the left inverse of block diagonal matrix $\Lambda(\hat{x})$ defined as:

$$\Lambda(\hat{x}) = \text{blockdiag} \left[I_{nk}, \frac{\partial f_k(x^{1:k}, u)}{\partial x^{k+1}}, \dots, \prod_{i=1}^{q-1} \frac{\partial f_k(\hat{x}, u)}{\partial \hat{x}^{k+1}} \right]$$

$$\Delta_\theta(\hat{x}) = \text{blockdiag} \left(I_{n_1}, \frac{1}{\theta} I_{n_1}, \dots, \frac{1}{\theta^{q-1}} I_{n_1} \right), \theta > 0$$

θ is a real number representing the only design parameter of the observer.

S is a definite positive matrix, solution of the following algebraic Lyapunov equation:

$$S + A^T S + SA = C^T C \quad (35)$$

With $C = [I_{n_1}, 0_{n_1}, \dots, 0_{n_1}]$ and $\mathcal{A} = \begin{bmatrix} 0 & \bar{\mathcal{A}} \\ 0 & 0 \end{bmatrix}$, with:

$$\bar{\mathcal{A}} = \text{blockdiag}(I_{n_1}, 0_{n_1}, \dots, 0_{n_1}) \in \mathbb{R}^{n_1(q-1)}$$

Note that relation (35) is independent of the system parameters and the solution can be expressed analytically. For a straightforward computation, its stationary solution is given by:

$$S(i, j) = (-1)^{i+j} C_{i+j-2}^{j-1} I_{n_1} \quad (36)$$

Where $C_j^i = \frac{j!}{i!(j-i)!}$ for $1 \leq i, j \leq q$.

In these conditions we can explicitly determinate the correction gain of (34) as follows:

$$\theta \Lambda^{-1}(\hat{x}) \Delta_\theta^{-1} S^{-1} C^T = \begin{bmatrix} \theta C_1^q I_{n_1} \\ \theta^2 C_2^q \left[\frac{\partial f_1}{\partial x^2}(x, u) \right]^- \\ \vdots \\ \theta^q C_q^q \left[\prod_{i=1}^{q-1} \frac{\partial f_k}{\partial x^{k+1}}(x, u) \right]^- \end{bmatrix} \quad (39)$$

It should be emphasized that the implementation of HGO is quite simple.

5. SIMULATION RESULTS AND COMMENTS

To investigate the usefulness of the proposed sensorless control approach a simulation experiments have been performed for a three-phase induction motor, whose parameters are depicted in Table 1.

Table 1. Induction motor parameters

Symbol	Quantity	N. Values
P_a	Power	0.75KW
F	Supply frequency	50HZ
P	Number of pair poles	2
V	Supply voltage	220V
R_s	Stator resistance	10 Ω
R_r	Rotor resistance	6.3 Ω
L_s	Stator inductance	0.4642H
L_r	Rotor inductance	0.4612H
L_m	Mutual inductance	0.4212H
ω_r	Rotor angular velocity	157rd/s
J	Inertia coefficient	0.02Kg ² /s
f	Friction coefficient	0N.s/rd

Two schemes of high gain state observer for the estimation of IM states are investigated. The first scheme is dedicated to electromagnetic state variables estimation of the considered induction motor, while the second scheme performs the estimation of mechanical state variables namely the rotor speed and the load torque.

Based upon the estimated and measured state variables, Backstepping controllers of the rotor speed and rotor flux are respectively implemented using Matlab/Simulink software programming. The obtained simulation results are presented in Figure 1 to Figure 5.

From Figure 1 to Figure 5 (Figure 1-5) the reference, measured and estimated state variables of the machine are presented according to load torque variation from no load value to the value $T_l = 5N.m$, introduced between [0.5s-1.5s]. This simulation is carried out by applying a reference speed as illustrated in Figure 1. The measured and estimated speed converges perfectly to their reference. One can see also that a significant decoupling effect of flux components under rotor angular speed and load torque variations. Figure 2 shows that measured and estimated rotor flux tracks the reference flux with no disturbance are found, Figure 3 note that the proposed approach exhibits high accuracy in torque tracking when the reference torque change, figures (4-5) show the measured and estimated stator currents and rotor flux components respectively. Analysis of the simulation results shows that the obtained performance of rotor angular speed and flux tracking are very adequate. Analysis of the different figures points out that designed nonlinear observer (High gain) effectively estimates the unmeasured state variables of the machine and tracks the load torque variations with respect to applied nonlinear control law computed in accordance with the Backstepping control technique.

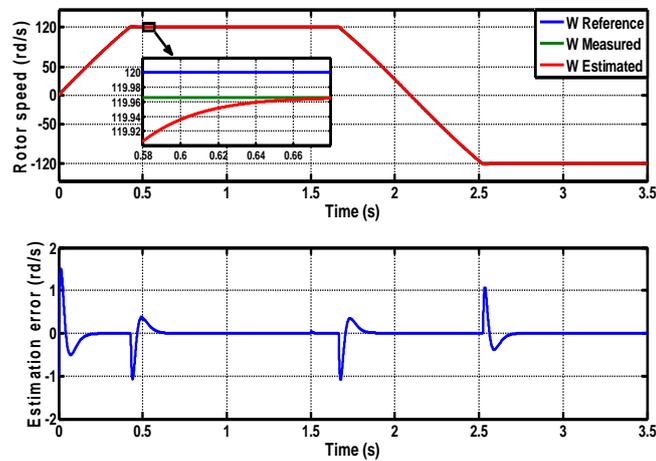


Figure 1. Reference measured and estimated Rotor speed evolution according to load variations

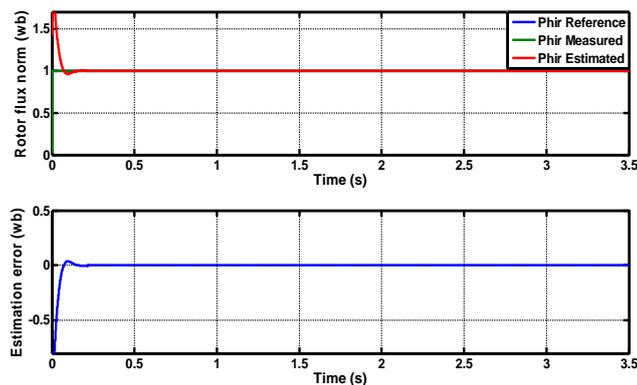


Figure 2. Reference measured and estimated norm of the rotor flux and estimation error

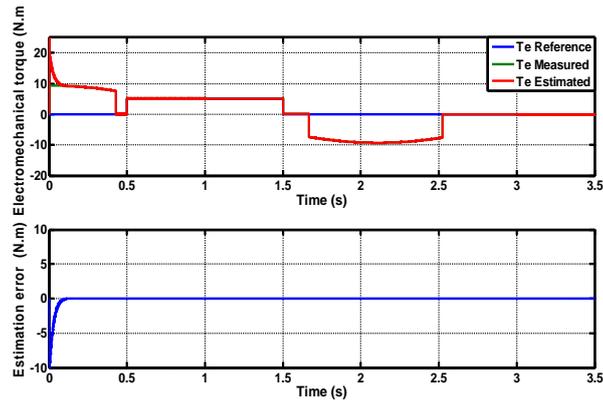
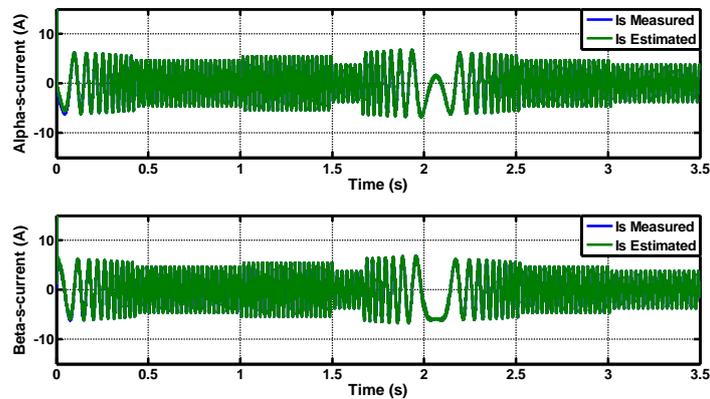
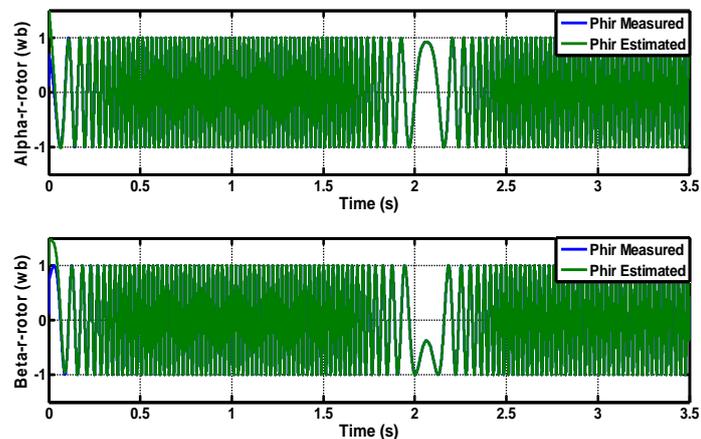


Figure 3. Reference measured and estimated electromechanical torque

Figure 4. Measured and estimated $(\alpha - \beta)$ stator currentsFigure 5. Measured and estimated $(\alpha - \beta)$ rotor flux

6. CONCLUSION

In this paper, we have investigate the possibility to implement a sensorless speed control of the induction machine using the technique of Backstepping, associated with a speed observer based on the high gain approach. The simulation results showed that this approach of control presents good performances and allows a complete decoupling between the flux and the torque. The machine keeps these performances. This technique can be improved further by using online estimation of parameters. On the other hand, simulation results show that this approach improves the performance of trajectory tracking and should bypass shortcomings of conventional methods. To this end, experimental tests will be investigated in a future framework.

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