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Full Length Research Paper

A robust nonlinear observer for states and parameters estimation and on-line adaptation of rotor time constant in sensorless induction motor drives

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This paper presents a robust non linear observer for variables and parameters estimation in sensorless Indirect Field Oriented Control (IFOC) of induction motors (IM). Based on the reduced order of the IM model, the rotor fluxes and time constant are estimated with the High Gain Observer (HGO) using only the stator currents and voltages. This reduced order model offers many advantages for real time identification and fault diagnosis of the IM. The major contributions of this work are: first, avoid the use of fluxes and speed sensors which increases the installation cost and degrades the mechanical robustness. Second, by reducing the order of the IM model, the implementation of the proposed observer doesn't require a very effective Digital Signal Processor (DSP). Finally, we show that the proposed control scheme is not sensitive to disturbances and parametric errors and it is robust against load variations and measurement noises. Simulation results are provided to prove the effectiveness of the proposed method.

Key words: Induction motor, field oriented control, sensorless, nonlinear observer, parameters estimation.

INTRODUCTION

In recent years significant advances have been made on the control of IM on the basis of rotor speed, flux and stator current measurements (Messaoudi, 2006). Tacho-generators or digital shaft-position encoders are usually used to detect the rotor speed of motors. Although, the fluxes of the IM can be directly measured by hall sensors or sensing coils. These speed and flux sensors caused many problems, such as degradation in mechanical robustness, increased cost, increased volume and lower the system reliability and require special attention to noise. In addition, for some special applications such as very high-speed motor drives and in hostile environment, there exist difficulties in mounting these sensors (Jeon et al., 2002). Therefore, flux and rotor speed are usually estimated with measured stator currents and voltages (Varghese and Sanders, 1998).

In the last decade, many researches have been carried on the design of sensorless control schemes of the IM. Most methods are basically based on the Model Reference Adaptive System schemes (MRAS) (Sbita and Ben Hamed, 2007a). In Sanchez et al. (2002) the authors used a reactive-power-based-reference model derived from Garcia-Correda and Robentsen (1999) in both motoring and generation modes but one of the disadvantages of this algorithm is its sensitivity to detuning in the stator and rotor inductances. The basic MRAS algorithm is very simple but the greatest drawback is the sensitivity to uncertainties in the motor parameters. Another method based on the Extended Kalman Filter (EKF) algorithm is used by Lee and Chen (1998), Messaoudi (2006), Messaoudi et al. (2007). The EKF is a stochastic state observer where nonlinear equations are linearized in every sampling period (Messaoudi, 2006). An interesting feature of the EKF is its ability to estimate simultaneously the states and the parameters of a dynamic process. This is generally useful for both the control and the diagnosis of the process. In Messaoudi

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et al. (2007) the authors used the EKF algorithm to simultaneously estimate variables and parameters of the IM in healthy case and under different faults, Lee and Blaabjerg (2006); Kwon et al. (2005) used the Luenberger Observer for state estimation of IM. The Extended Luenberger Observer (ELO) is a deterministic observer which also linearizes the equations in every sampling period. Another type of methods for state estimation which is based on the intelligent technique is used in the recent years by many authors (Kenne et al., 2006; Sbita and Ben Hamed, 2007c). Fuzzy logic and neural networks has been a subject of growing interest in recent years. Neural network and fuzzy logic algorithms are quite heavy for basic microprocessors. In addition, several papers provide sensorless control of IM which are based on the variable structure technique (Kim et al., 2004) and the HGO (Besancon et al., 2002; Abdellah et al., 2004) which is a powerful observer that can estimate simultaneously variables and parameters of a large class of a nonlinear systems and doesn't require a high performance processor for real time implementation. Now, the IM is widely used in many industrial applications due to its reliability, ruggedness, and relatively low cost. Thanks to the advances of power electronics and DSP technology, the control schemes of the IM are developed from simple scalar control methods to sensorless or auto-tuning control strategies (FOC, DTC and Control via input-output linearization,...) (Bodson et al., 1994; Sbita and Ben Hamed, 2007b).

In this paper, we present the design of a high performance sensorless IM vector control which is robust against rotor time-constant variations. These variations are predicted and corrected by using the HGO. This non-linear observer is developed to simultaneously estimate the mechanical speed, the rotor fluxes and the rotor time-constant. Simulation results are presented to highlight the effectiveness and the robustness of the proposed control scheme against measurement noises, rotor time-constant and load torque variations.

Induction motor model

The IM mathematical model, in space vector notation, established in d-q coordinate system rotating at synchronous speed ω_s is given by the following equations:

Electric equations:

$$\begin{aligned} v_{ds} &= R_s i_{ds} + \frac{d\varphi_{ds}}{dt} - \omega_s \varphi_{qs} \\ v_{qs} &= R_s i_{qs} + \frac{d\varphi_{qs}}{dt} + \omega_s \varphi_{ds} \\ 0 &= R_r i_{dr} + \frac{d\varphi_{dr}}{dt} - \omega_{sl} \varphi_{qr} \\ 0 &= R_r i_{dr} + \frac{d\varphi_{qr}}{dt} + \omega_{sl} \varphi_{dr} \end{aligned} \quad (1)$$

Magnetic equations:

$$\begin{aligned} \varphi_{ds} &= L_s . i_{ds} + M . i_{dr} \\ \varphi_{qs} &= L_s . i_{qs} + M . i_{qr} \\ \varphi_{dr} &= L_r . i_{dr} + M . i_{ds} \\ \varphi_{qr} &= L_r . i_{qr} + M . i_{qs} \end{aligned} \quad (2)$$

Electromagnetic torque equation:

$$T_e = p \frac{M}{L_r} \left(\varphi_{dr} i_{qs} - \varphi_{qr} i_{ds} \right) \quad (3)$$

Mechanical equation:

$$\dot{\omega}_r = \frac{p^2 M}{J L_r} \left(\varphi_{dr} i_{qs} - \varphi_{qr} i_{ds} \right) - \frac{p}{J} T_l - \frac{f}{J} \omega_r \quad (4)$$

Combining equations (1)-(4), we can write the motor model as:

$$\begin{bmatrix} \dot{i}_{ds} \\ \dot{i}_{qs} \\ \dot{\varphi}_{dr} \\ \dot{\varphi}_{qr} \\ \dot{\omega}_r \end{bmatrix} = \begin{bmatrix} -\gamma i_{ds} + \omega_s i_{qs} + \frac{k}{T_r} \varphi_{dr} + k \omega_r \varphi_{qr} + \frac{1}{\sigma L_s} v_{ds} \\ -\omega_s i_{ds} - \gamma i_{qs} - k \omega_r \varphi_{dr} + \frac{k}{T_r} \varphi_{qr} + \frac{1}{\sigma L_s} v_{qs} \\ \frac{M}{T_r} i_{ds} - \frac{\varphi_{dr}}{T_r} + \omega_{sl} \varphi_{qr} \\ \frac{M}{T_r} i_{qs} - \omega_{sl} \varphi_{dr} - \frac{\varphi_{qr}}{T_r} \\ \frac{M}{L_r} \frac{p^2}{J} \left(\varphi_{dr} i_{qs} - \varphi_{qr} i_{ds} \right) - \frac{f \omega_r}{J} - \frac{p T_l}{J} \end{bmatrix} \quad (5)$$

Where:

$\gamma = \left(\frac{R_s}{L_s} + \frac{R_r (1 - \sigma)}{L_r \sigma} \right)$, $k = \frac{(1 - \sigma)}{M \sigma}$ and $\sigma = 1 - \frac{M^2}{L_s L_r}$.

Rotor field orientation

There are many categories of vector control strategies. We are interested in this study to the so-called IFOC. We have shown in equation (3) that the electromagnetic torque expression, in the dynamic regime, presents a coupling between stator current and rotor flux. The main objective of the vector control of IM is, as in direct current (DC) machines, to independently control the torque and the flux. This is done by using a d-q rotating reference frame synchronously with the rotor flux space vector. The d-axis is aligned with the rotor flux space vector. Under this condition we have; $\varphi_{dr} = \varphi_r$ and $\varphi_{qr} = 0$.
In this case the torque equation becomes:

$$T_e = p \frac{M}{L_r} \varphi_r i_{qs} \quad (6)$$

It is right to adjust the flux while acting on the component i_{ds} of the stator current and to adjust the torque while acting on the i_{qs} component. One has two variables of action then as in the case of a DC machine.

Combining equations (1) and (2) we obtain the following d and q-axis stator currents:

$$i_{ds} = \frac{(1+T_r s)}{M}\varphi_r^* \tag{7}$$

$$i_{qs} = \frac{T_r}{M}\omega_{sl}^* \varphi_r^* \tag{8}$$

We replace i_{qs} by its expression to obtain the torque T_e as a function of the reference slip speed.

$$T_e = p\frac{\varphi_r^{*2}}{R_r}\omega_{sl}^* \tag{9}$$

Using equations (1)-(2) we obtain the following voltage equations:

$$v_{ds} = (R_s + \sigma L_s s) i_{ds} + \frac{M}{L_r} p\varphi_r^* - \sigma L_s \omega_s i_{qs} \tag{10}$$

$$v_{qs} = (R_s + \sigma L_s s) i_{qs} + \sigma L_s \omega_s i_{ds} + \frac{M}{L_r}\omega_s \varphi_r^* \tag{11}$$

These equations are functions of some structural electric parameters of the IM (R_s, R_r, L_s, L_r, L_m), which are in reality approximate values. We will come back thereafter to the influence of the bad knowledge of most interest parameter ($T_r = L_r / R_r$) on the control of the machine (Alamir, 2002; Mastorocostas et al., 2006).
The rotor flux amplitude is obtained by solving (8), and its spatial position is given by:

$$\theta_s = \int\left(\omega_r + \frac{M i_{qs}}{T_r \varphi_r}\right)dt \tag{12}$$

High gain observer

Consider the nonlinear system:

$$\begin{cases} \dot{x} = f(x) + g(x)u \\ y = h(x) \end{cases} \tag{13}$$

Where, $x \in \mathbb{R}^n$ is the state, $u \in \mathbb{R}^m$ is the input and $y \in \mathbb{R}^l$ is the output.

The system (13) must be uniformly observable. Then, it is possible to apply a change of variable $z = \Gamma(x)$ which will transform the system (13) to the following forms:

$$\begin{cases} \dot{z} = Az + \varphi(u, z) \\ y = Cz \end{cases} \tag{14}$$

Using some Lipschitz property of φ in x uniformly in u, an exponential observer for (14) can be designed as follows (Gauthier and Bornard, 1992):

$$\dot{\hat{z}} = A\hat{z} + \varphi(\hat{z}, u) - S_\theta^{-1} K (C\hat{z} - y) \tag{15}$$

Where K is such that A - KC is stable, S_θ is the solution of the Lyapunov equation:

$$\dot{S}_\theta = -\theta S_\theta - A^T S_\theta - S_\theta A + C^T C = 0 \tag{16}$$

θ is a positive and commonly sufficient large design parameter (Hammouri and Marchand, 1991; Moreno and Vargas, 2000). By applying an inverse change of variable to come back to the initial non-linear system, the observer for the system (13) is given by:

$$\dot{\hat{x}} = f(\hat{x}) + g(\hat{x})u - \left(\frac{\partial \Gamma}{\partial \hat{x}}(\hat{x}(t))\right)^{-1} S_\theta^{-1} C^T (h(\hat{x}) - y) \tag{17}$$

The demonstration is made in Bornard and Hammouri (1991). Γ is an $\mathbb{R}^n \rightarrow \mathbb{R}^n$ application

$$\Gamma = \left[h_1, L_f h_1, \cdots, L_f^{\delta_1} h_1, \ldots, h_p, L_f h_p, \ldots, L_f^{\delta_p} h_p\right]^T \tag{18}$$

Let $L_f^d h$ denotes the Lie derivative of the scalar function h with respect to the vector field f (Kocarev et al., 1998):

$$L_f h = \frac{\partial h}{\partial x} f(x) = \sum_{i=1}^{n} \frac{\partial h}{\partial x_i} f_i(x) \tag{19}$$

Recursively, we have

$$L_f^{\delta} h = L_f\left(L_f^{\delta-1} h\right) \tag{20}$$

The exponential convergence of this observer is theoretically proved by Besançon et al. (2002). The most difficult stage in the synthesis of the HGO is the calculation of $S_\theta^{-1}(\theta)$ and $\left(\frac{\partial \Gamma}{\partial \hat{x}}(\hat{x}(t))\right)^{-1}$.

Application to the induction motor

It is well known that most control systems for IM require the knowledge of the rotor flux as well as that of the angular speed (Sbita and Ben Hamed, 2007a).

The IM model is given by the equation (5) where:

$$x = \left[i_{ds}\ i_{qs}\ \varphi_{dr}\ \varphi_{qr}\ \omega_r\right]^T,$$

$$y = \left[i_{ds}\ i_{qs}\right]^T, \quad u = \left[v_{ds}\ v_{qs}\ T_l\right]^T.$$

$$f(x) = \begin{bmatrix} -\gamma i_{ds} + \omega_s i_{qs} + \frac{k}{T_r}\varphi_{dr} + k\omega_r\varphi_{qr} \\ -\omega_s i_{ds} - \gamma i_{qs} - k\omega_r\varphi_{dr} + \frac{k}{T_r}\varphi_{qr} \\ \frac{M}{T_r} i_{ds} - \frac{\varphi_{dr}}{T_r} + \omega_{sl}\varphi_{qr} \\ \frac{M}{T_r} i_{qs} - \omega_{sl}\varphi_{dr} - \frac{\varphi_{qr}}{T_r} \\ \frac{M}{L_r}\frac{p^2}{J}\left(\varphi_{dr} i_{qs} - \varphi_{qr} i_{ds}\right) - \frac{f\omega_r}{J} \end{bmatrix};$$

$$g(x) = \begin{bmatrix} \frac{1}{\sigma L_s} \\ \frac{1}{\sigma L_s} \\ 0 \\ 0 \\ -\frac{p}{J} \end{bmatrix}; h(x) = \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix}. \quad (21)$$

Solving equations (16) - (18) leads to:

$$\frac{\partial \Gamma}{\partial t}(x) = \begin{pmatrix} 1 & 0 & 0 & 0 & 0 \\ -\gamma & w_s & \frac{k}{T_r} & k\hat{w}_r & k\hat{\varphi}_{qr} \\ 0 & 1 & 0 & 0 & 0 \\ -w_s & -\gamma & -k\hat{w}_r & \frac{k}{T_r} & -k\hat{\varphi}_{dr} \\ 0 & 0 & 0 & 0 & 1 \end{pmatrix}; \quad (22)$$

$$S_\theta^{-1}(\theta)C^T = \begin{pmatrix} 2\theta & 0 \\ 0 & 2\theta \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \end{pmatrix}. \quad (23)$$

Rotor time-constant estimation

Because it is highly sensitive to parameters variation especially, the rotor time-constant, which is mainly a function of temperature (Huai et al., 2003; Mastorocostas et al., 2006); hence, we are interest in this section to the estimation of this parameter.

The model described by equation (5) has many disadvantages for state or parameter estimation. First, because two state components (the stator currents) of this full-order model are easily and accurately measurable, full-order observers deduced from this model are needlessly complicated and implies a small sampling period 1 ms or less. Second, the model contains simultaneously fast and slow modes: the current dynamics is far higher than the rotor flux dynamics (Zein et al., 2001). For this reason it is quite necessary to reduce the observer order to decrease computation and to avoid small sampling period. Thus, the experimental implementation of this observer becomes easier to do.

To get a simple model for rotor flux and parameter estimation, we will remove the stator currents which can be easily measured, from the state vector. Then we consider the stator currents as input: $u = [i_{ds}\ i_{qs}]^T$ and the rotor flux, estimated in an open loop using stator voltage equations, as output: $y = [\varphi_{dre}\ \varphi_{dre}]^T$. Then, the state vector only consists of the rotor flux components and the inverse of the rotor time-constant β_r: $x = [\varphi_{dr}\ \varphi_{dr}\ \beta_r]^T$.
In this case:

$$\frac{\partial \Gamma}{\partial t}(x) = \begin{pmatrix} 1 & 0 & 0 \\ -\hat{\beta}_r & w_{sl} & -\left(\hat{\varphi}_{dr} - M i_{ds}\right) \\ 0 & 1 & 0 \end{pmatrix} \quad (24)$$

$$S_\theta^{-1}(\theta)C^T = \begin{pmatrix} 2\theta & 0 \\ 0 & 2\theta \\ 0 & 0 \end{pmatrix} \quad (25)$$

Open loop observer

The rotor flux can be calculated by using the stator equation of the IM.

$$\begin{cases} \dot{\varphi}_{dre} = \frac{L_r}{M}\left(v_{ds} - R_s i_{ds}\right) - \frac{\sigma L_s L_r}{M}\left(s i_{ds} - \omega_s i_{qs}\right) + \omega_s \varphi_{qre} \\ \dot{\varphi}_{qre} = \frac{L_r}{M}\left(v_{qs} - R_s i_{qs}\right) - \frac{\sigma L_s L_r}{M}\left(s i_{qs} - \omega_s i_{ds}\right) + \omega_s \varphi_{dre} \end{cases} \quad (26)$$

This expression is also known as the *voltage model* of the machine. It presents similar problems and advantages, since it requires a pure integration for flux estimation and it is sensitive to errors in the stator resistance at low speed. This makes the voltage model unsuitable for low speed operation (Abdellah et al., 2004).

The fact that the state vector only consists of the rotor flux and the rotor time-constant offers a double advantage. First, the reduction of the state dimension shortens the computational volume and complexity. Second, thanks to its low dynamic, the rotor flux can be easily estimated. If the rotor time-constant is sufficiently accurate, though, IRFOC IM offers a good decoupling between flux and torque. Therefore, we obtain faster and more accurate torque control and greater stability.

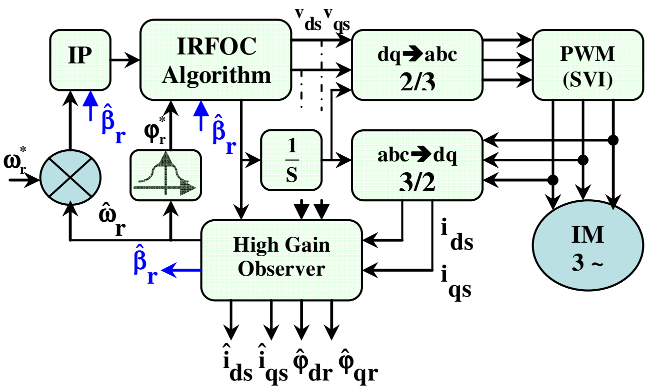

Figure 1. Block diagram of the sensorless IRFOC IM drive system.

Table 1. Motor data

Rated power	3 kW
Rated speed	1440 rpm
frequency	50 Hz
Voltage (Δ / Y)	220 V / 380 V
Voltage (Δ / Y)	11.5 A / 6.6 A
p	2
R_s	2.3 Ω
R_r	1.55 Ω
$L_s = L_r$	0.261 H
M	0.249 H
J	0.0076 kg.m^2

For speed controller (Figure 1), we have designed an integral and proportional (IP) speed controller in order to stabilize the speed-control loop. The gains of the IP controller, *kp* and *ki*, are determined using a design method to obtain a damping ratio of 1 (Messaoudi, 2006). The obtained gain values of the IP speed controller are:

$$k_p = \frac{J - \tau_1 f}{p\,\tau_1\,\lambda_p} \quad (27)$$

$$k_i = \frac{p\,k_p\,\lambda_p + f}{4 p k_p \lambda_p \tau_1} \quad (28)$$

Where: $\tau_1 = \dfrac{J}{p\,k_p\,\lambda_p + f}$ and $\lambda_p = p\dfrac{T_r}{L_r}\varphi_r^*$

These gains, as shown by (27) and (28), depend on the rotor time-constant. This rotor time-constant is continuously adjusted to its actual value estimated by the implemented HGO. With this adaptation, which is made on-line, the proposed IRFOC scheme is robust against rotor time-constant deviations and ensures the decoupleing process to be continuously achieved (Messaoudi et al., 2007).

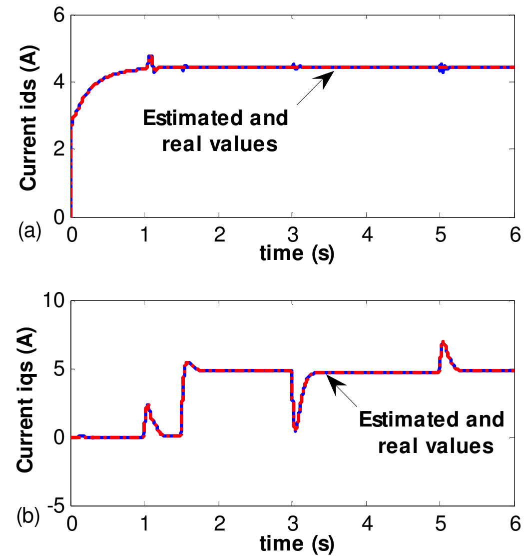

Figure 2. Actual and estimated stator currents, (a) direct components, (b) quadratic components.

RESULTS AND DISCUSSION

Simulations, using MATLAB Software Package, have been carried out to verify the effectiveness of the proposed method. The application of the HGO for unmeasured variables estimation is illustrated by a computer simulation shown in the block diagram of Figure 1.

The motor used in simulation data are given in Table 1. Flux reference is set to its rated value of 1.1 Wb and speed reference is set from 0 to 157 rad/s at t=1s, and the load torque is varied 3 times; at 2 s from 0 to 5 Nm, at 3.5 s from 5 to the rated load 10 Nm and at 5 s is increased to 3 Nm. The initial state vector is set to zero, unless B_r is initialized to 5.8 a value near its known rated value to minimise the estimation error at start-up. The sampling time is set to 1 ms.

Figures 2 and 3 show respectively the real and the estimated stator currents and rotor fluxes. Figure 4 shows the load and the electromagnetic torque. Figure 5 illustrates the real and the estimated speed under different load conditions

By solving the Lyapunov equation (16) we find the matrix S_θ which is function of θ. Hence, the dynamic performances of the HGO is realised by only tuning the parameter θ which called gain. This gain is chosen in order to assure the desired performances.

i) A high value of θ decreases the observer time convergence, but also it leads to a high oscillation and a

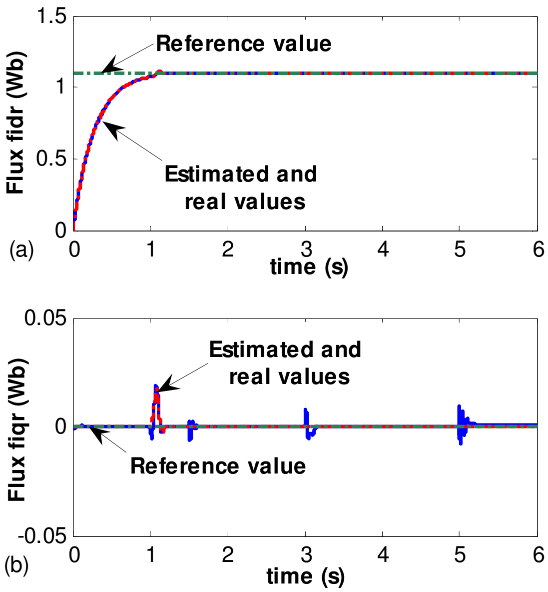

Figure 3. Actual and estimated rotor flux, (a) direct components, (b) quadratic components.

large estimation error at the start-up,
ii) A low value of θ reduces the oscillations and the estimation error at the start-up, but it increases the observer time convergence.

In fact, it is right to choose an optimal value of θ (here it is fixed to θ =1.86) which fulfil the desired performances.

Figure 6 gives the actual and the estimated rotor time-constant under heavily loaded condition (nominal torque).

Figure 7 shows the actual and estimated rotor fluxes. When the inverse of the rotor time-constant increased 100% at 2 s from its rated value, the decoupling is lost and the control performance is degraded. But the HGO keeps its effectiveness and the estimated variables converge to their real values.

Figure 8 gives the actual and the estimated rotor time-constant. Thus, the rotor resistance is stepped up 100% at 2 s from its rated value. This variation is predicted and corrected by using the HGO. The on-line estimated rotor time-constant actual value is fed forward to the controller and to the slip speed calculation module to obtain robust control performance with the proposed IFOC scheme.

Figure 9 shows the actual and estimated rotor fluxes with on-line adaptation of the IFOC equations and the IP speed controller gains with the estimated rotor time-constant. With this adaptation, which is made on-line, the proposed vector control scheme is robust against rotor time-constant deviations and ensures the decoupling process to be continuously achieved.

The reversal speed response of the IM is shown in Figures 10 to 13 under 10% of rated load. The reference speed is set to 4 rad/s at t = 4 s. Then the set point is

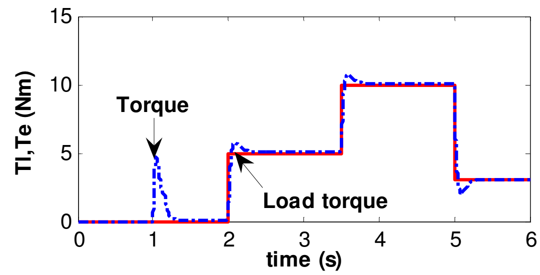

Figure 4. Electromagnetic torque and load torque.

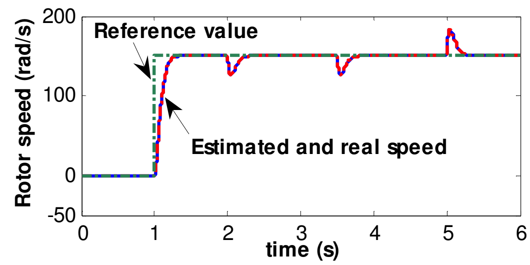

Figure 5. Actual and estimated rotor speed.

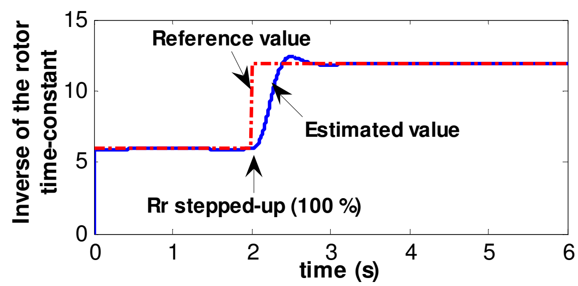

Figure 6. Actual and estimated rotor time-constant.

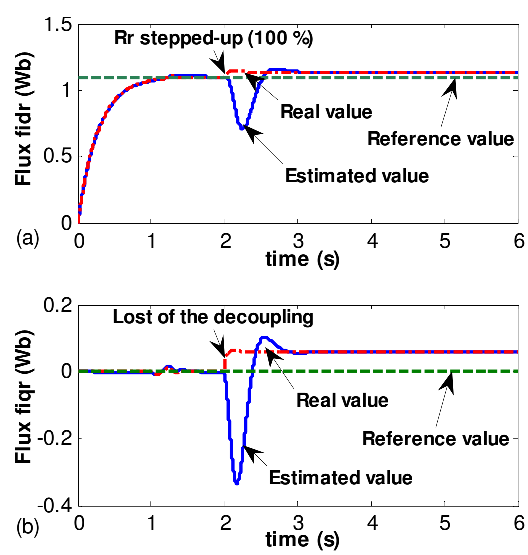

Figure 7. Actual and estimated rotor flux, (a) direct components, (b) quadratic components.

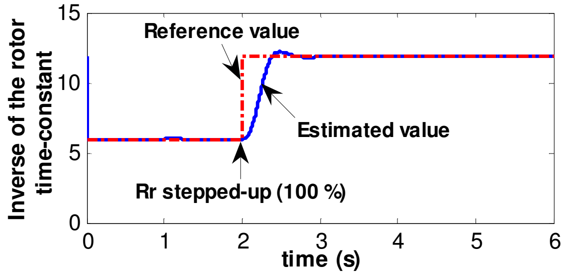

Figure 8. Actual and estimated rotor time-constant with IRFOC adaptation.

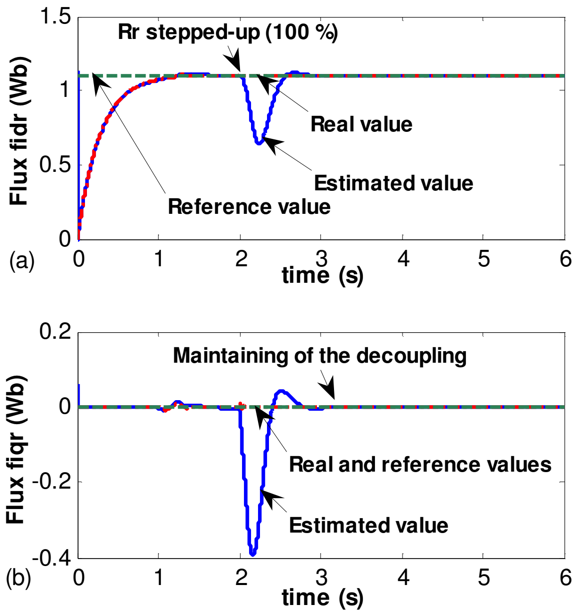

Figure 9. Actual and estimated rotor fluxes with IRFOC adaptation, (a) direct component, (b) quadratic component.

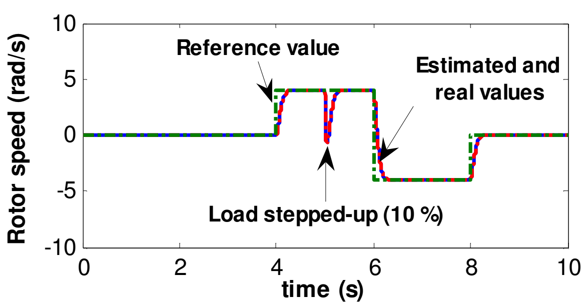

Figure 10. Actual and estimated rotor speed at low speed region.

changed to -4 rad/s at t = 6 s. These figures show the dynamic behaviour of the IM at low speeds. The results show that the IRFOC has a good tracking performance even at low speeds.

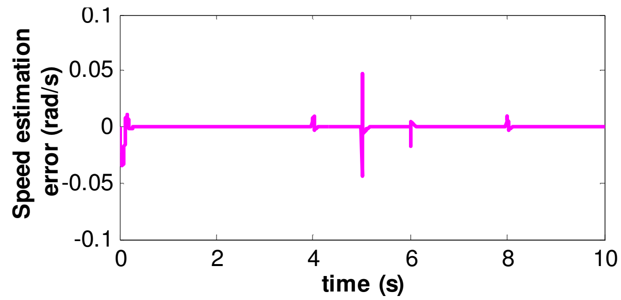

Figure 11. Speed estimation error at low speed region.

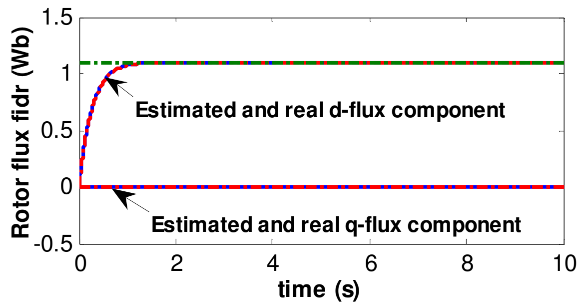

Figure 12. Actual and estimated rotor fluxes at low speed region (direct and quadratic components).

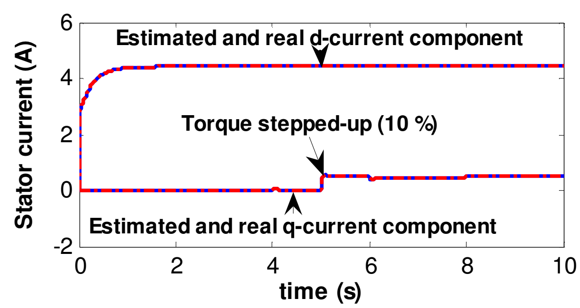

Figure 13. Actual and estimated stator currents at low speed region (direct and quadratic components).

Observation results in presence of measurement noises

To highlight the robustness of the observer, white Gaussian noises with variances of 10^{-2} are simultaneously added to the measured stator currents and voltages. Figure 14 shows the real an estimated d and q components of the rotor fluxes. The real and estimated inverse of the rotor time constant is given in Figure 15.

It clearly appears that the HGO preserves its performances and robustness in presence of noises. The on line estimation of the IM states and parameters is tested by many researchers and is proved to give satisfactory results. The most used techniques to estimate these states and parameters are the EKF (Messaoudi et al.,

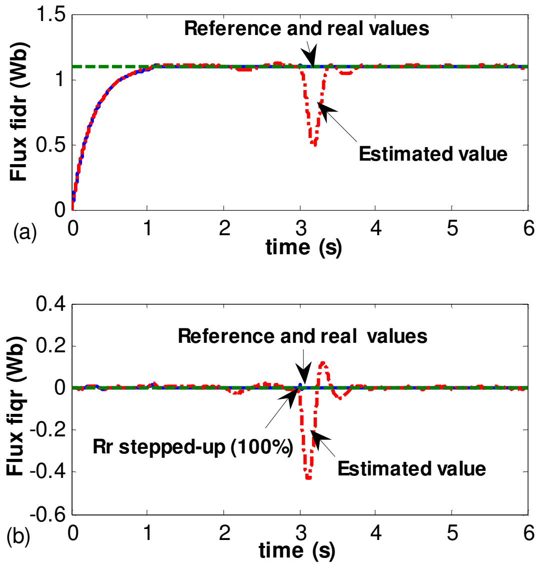

Figure 14. Real and estimated rotor flux in presence of measure noises with IRFOC adaptation, (a) direct components, (b) quadratic components.

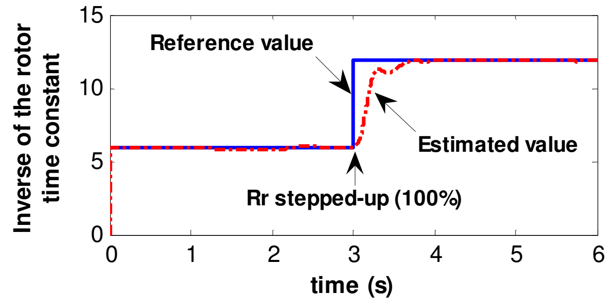

Figure 15. Real and estimated rotor time-constant in presence of measure noises.

2007; Ouahrouche, 2000), the MRAS (Beguenane and Ouhrouche, 2003; Sbita and Ben Hamed, 2007a) and the Luenberger observer (Ben Hamed et al., 2006). The contribution of the work presented in this paper is the design of an indirect vector control of the IM which is robust against rotor time constant, load variations and measurement noises using a reduced order HGO. This observer is less difficult in experimental implementation than the EKF and the Luenberger observer, which required intensive computations due to numerous matrixes manipulation, and more effective than the MRAS, which can not estimate simultaneously states and parameters.

Conclusion

In this paper, a new adaptive approach for IFOC induction motor drive has been proposed. Simulation results confirm the effectiveness of the proposed HGO, which allows one to achieve an efficient rotor time-constant tracking and to avoid the drawbacks of the use of sensor sets in the control scheme of the IM. The estimated rotor time-constant is fed forward to the speed controller and to the slip speed calculation module to obtain robust control performance with the proposed IFOC scheme.

In addition, the simulation results show the robustness of the proposed algorithm against measurement noises, rotor time-constant and load torque variations. Also, it has shown that the HGO gives good results at low speed region, including zero speed operation.

In a future fellow up work, the proposed scheme is to be implemented on a DSP based on the 16 bits floating point arithmetic Texas Instrument TMS320C31 processor.

Table 2. Nomenclature

v_{ds}, v_{qs}	d- and q-axis stator voltages
i_{ds}, i_{qs}	d- and q-axis stator currents
i_{dr}, i_{qr}	d- and q-axis rotor currents
$\varphi_{ds}, \varphi_{qs}$	d- and q-axis stator flux linkages
$\varphi_{dr}, \varphi_{qr}$	d- and q-axis rotor flux linkages
R_s, R_r	stator and rotor resistances
L_s, L_r	stator and rotor inductances
M	mutual inductance
w_r	rotor angular speed
w_s	synchronous angular speed
w_{sl}	slip angular speed
σ	total leakage coefficient
T_e, T_l	electromagnetic and load torque
J	total inertia
f	viscous friction coefficient
p	number of pole pairs
^	denotes the estimated value

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