Real-time Estimation of Rotor Time Constant and Flux of an Induction Motor: Experimental Results

1O. Asseu, 1A. Tanoh, 1M. Koffi, 1Z. Yeo, 1T.R. Ori and 2X. Lin-Shi
1Department of Electrical and Electronic Engineering, Institute National Polytechnique Houphouët Boigny, BP 1093 Yamoussoukro Côte d'Ivoire
2Université de Lyon, AMPERE, INSA Lyon, CNRS UMR 5005, Villeurbanne 69621, France

Abstract: In this study, a reduced-order discrete-time Extended Sliding Mode Observer (ESMO) is introduced for on-line estimation of rotor fluxes and rotor time constant. The satisfying simulations and experimental results carried out on a 5.5 kW induction motor demonstrate the excellent performance and high robustness of the proposed ESMO against parameter variations, modeling uncertainty and measurement noise. It is concluded that the implementation in real-time of this reduced-order ESMO in the industrial applications, compared with the full-order ESMO, permit on the one hand to overcome the heavy computational effort, complexity and hard tuning of the estimation algorithm and on the other hand, to reduce the execution time of the observation with a good accuracy and considerable rapidity.

Keywords: Induction motors, field oriented control, sliding mode observer, test bench

INTRODUCTION

The Induction Motors (IM) is widely used in industrial applications due to its reasonable cost, robust qualities and simple maintenance. However, the control of IM drives is proved very difficult since the dynamic systems are non linear, the electric rotor variables (such as flux, torque) are not measurable and the physical parameters are often imprecisely known or variable. For instance, the rotor resistance drifts with the temperature of the rotor current frequency. One of the most significant developments in this area has been the Field-Oriented Control (FOC) (Montanari et al., 2000; Roncero-Sanchez et al., 2007) which allows an efficient control of the torque dynamics of an IM. However, a variation of the rotor resistance can induce a lack of field orientation. In order to achieve better dynamic performance, an on-line estimation of rotor fluxes and rotor resistance is necessary. In (Derdiyok, 2005, Amulili and Ali, 2007), a full-order Sliding Mode Observer (SMO) has been used with success for rotor fluxes estimation. This full-order SMO, built from the dynamic model of the IM by adding corrector gains with switching terms (Jingchuan et al., 2005; Benchiba et al., 1999; Shnai et al., 2007), is used to provide not only the unmeasurable state variable estimation (rotor fluxes) but also the estimation of the measurable parameters (stator currents). However, the determination of the measurable parameters estimation imposes some estimation algorithms very long and usually sophisticated with an increase

Corresponding Author: Zie Yeo, INP-HB, BP 1093 Yamoussoukro, Côte d'Ivoire
Tel: (+225) 05 01 71 35

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of the computational volume. Therefore, in order to reduce the computation rate of the estimation algorithms, the currents estimation is not necessary since they are measured.

So, after a brief review of the IM model and the conventional sliding mode observer, the main aim of this study is to present the experimental results carried out on a 5.5 kW IM drive system (Mendes et al., 2002) in order to show the effectiveness of the proposed Reduced-order Discrete-time Extended Sliding Mode Observer (RDESMO) method which permits to solve only the problem of the rotor fluxes and rotor time constant estimations.

**MATERIALS AND METHODS**

**Induction Motor Model and Full-Order Sliding Mode Observer**

This study, conducted in the Laboratory of applied Electrical and Electronic (INPHB Yamoussoukro, Côte d'Ivoire) from July 2006 to June 2009 by a theoretical work, has been implemented and validated in real-time on a test bench (Fig. 1), which is constructed and assembled in the research centre of Electrical Engineering, Institute National of Applied Sciences (INSA) in Lyon-France. Globally the test bench is composed of an induction motor, a powder brake completed by current and voltage sensors. An engine bench description is basically detailed in the results section.

By assuming that the saturation of the magnetic parts and the hysteresis phenomenon are neglected, the classical dynamic model of the induction motor in a (d, q) synchronous reference frame can be described by De Fornel and Louis (2007):

\[
\begin{align*}
V_a &= R_s I_a + \frac{d\Phi_a}{dt} + \omega_c \Phi_w \\
V_q &= R_s I_q + \frac{d\Phi_q}{dt} + \omega_c \Phi_w \\
\Phi_a &= L_v I_a + \sigma L_s I_a \\
\Phi_q &= L_v I_q + \sigma L_s I_q
\end{align*}
\]  

(1a)

\[
\begin{bmatrix}
\Phi_a \\
\Phi_q
\end{bmatrix} =
\begin{bmatrix}
L_v & L_v \\
L_v & L_v
\end{bmatrix}
\begin{bmatrix}
I_a \\
I_q
\end{bmatrix}
\]  

(1b)

From (1a) and (1b), using as state variables the rotor fluxes (\(\Phi_a, \Phi_q\)) and the stator currents (\(I_a, I_q\)), the electrical dynamic model of the IM, completed with the output equation, can be represented as a time-varying fourth-order system given by:

\[
x = f(x) + g(u), \quad y = [y_1, y_2] = [I_a, I_q]^T
\]

with

\[
x = [\Phi_a, \Phi_q, I_a, I_q]^T, \quad u = [V_a, V_q]^T
\]

\[
\text{Fig. 1: A global view of the test bench}
\]
\[
\begin{align*}
\mathbf{f}(x) &= \begin{bmatrix}
-\sigma, \Phi_s + \omega_q \Phi_y + \sigma_i L_a I_a \\
-\omega_q \Phi_s - \sigma \Phi_y + \sigma_i L_a I_a \\
\sigma \beta \Phi_s + \beta \omega_q \Phi_y - \lambda I_b + \omega_1 I_y \\
-\beta \omega_q \Phi_s + \beta \sigma \Phi_y - \omega_1 I_y - \lambda I_y
\end{bmatrix}, \\
\mathbf{g} &= \begin{bmatrix}
0 \\
0 \\
1/\sigma L_a \\
0
\end{bmatrix}, \\
\sigma &= \frac{1}{L_q}, \\
\lambda &= \lambda(\sigma_i) = \frac{1}{\sigma} \left( \frac{1}{L_q} + (1-\sigma) \sigma_i \right), \\
\beta &= \frac{1-\sigma}{\sigma L_a}, \\
\sigma &= \frac{L_a}{L_q L_i}.
\end{align*}
\] (2)

Moreover, by choosing a rotating reference frame (d, q) so that the direction of axe d is always coincident with the direction of the rotor flux representative vector (field orientation), it is well known that this rotor field orientation in a rotating synchronous reference frame realizes:

\[
\Phi_d = \Phi_q = \text{Constant and } \Phi_y = 0
\] (3)

Assume that among the state variables, only the stator currents noted \( z_1 \) and \( z_2 \) are measurable. Denote \( \hat{x}_1 \) and \( \hat{x}_2 \) the estimates of the fluxes \( \Phi_s \) and \( \Phi_y \). Consider that \( \hat{z}_1 \) and \( \hat{z}_2 \) the estimates of the stator currents \( I_a \) and \( I_y \). The classical full-order sliding mode observer is a copy of the model (2) by adding corrector gains with switching terms (Asseu et al., 2008):

\[
\begin{align*}
\dot{\hat{x}}_1 &= -\sigma \hat{x}_1 + \omega_q \hat{x}_2 + \sigma_i L_a z_1 + \Gamma_1 I_a, \\
\dot{\hat{x}}_2 &= -\omega_q \hat{x}_1 - \omega_1 \hat{x}_2 + \sigma_i L_a z_2 + \Gamma_1 I_y, \\
\dot{\hat{z}}_1 &= \beta \sigma \hat{x}_1 + \beta \omega_q \hat{x}_2 - \lambda z_1 + \omega_1 z_2 + \frac{1}{s} V_a + \Lambda_1 I_a, \\
\dot{\hat{z}}_2 &= -\beta \omega_q \hat{x}_1 + \beta \sigma \hat{x}_2 - \omega_1 z_1 - \lambda z_2 + \frac{1}{s} V_y + \Lambda_1 I_y
\end{align*}
\] (4)

where, \( \Gamma_1 \), \( \Gamma_2 \) and \( \Lambda_1 \), \( \Lambda_2 \) are the observer gains. The switching \( I_a \), that depends on the estimated currents, is given by:

\[
I_a = \begin{bmatrix}
\text{sign}(\hat{x}_1) \\
\text{sign}(\hat{x}_2)
\end{bmatrix}
\] (5)

Setting \( \hat{x} = x - \hat{x} \), \( \hat{z} = z - \hat{z} \), the estimation error dynamics is defined by:

\[
\begin{align*}
\dot{\hat{x}}_1 &= -\sigma \hat{x}_1 + \omega_q \hat{x}_2 - \Gamma_1 I_a, \\
\dot{\hat{x}}_2 &= -\omega_q \hat{x}_1 - \omega_1 \hat{x}_2 - \Gamma_1 I_y, \\
\dot{\hat{z}}_1 &= -\beta \sigma \hat{x}_1 + \beta \omega_q \hat{x}_2 - \lambda z_1 - \omega_1 z_2, \\
\dot{\hat{z}}_2 &= -\beta \omega_q \hat{x}_1 + \beta \sigma \hat{x}_2 - \omega_1 z_1 - \lambda z_2, \\
\end{align*}
\]

The condition for convergence is verified by chosen the following observer gain matrices:

\[
A_1 = \begin{bmatrix}
\beta \sigma & \beta \omega_q \\
-\beta \omega_q & \beta \sigma
\end{bmatrix}, \\
A_2 = \begin{bmatrix}
\Gamma_1 \\
\Gamma_2
\end{bmatrix}, \\
\Delta = \begin{bmatrix}
q - \sigma & \omega_q \\
-q - \sigma & q
\end{bmatrix}, \\
\Lambda = \begin{bmatrix}
n & 0 \\
0 & n
\end{bmatrix}
\] (6)

where, \( q \) and \( n \) are two positive adjusting parameters.


Reduced-order Discrete-Time Extended Sliding Mode Observer

As previously underlined, a variation of the rotor resistance can induce performance degradation of the system. Also the stator currents are measurable therefore their on-line estimation is not necessary. Thus a Reduced-order Extended Sliding Mode Observer (RESMO) is introduced. In order to estimate the rotor flux and rotor time constant \( (\Omega_r = 1/T_r = -R/\Lambda) \) variations, a three-dimensional state vector defined by \( X_r = [\Phi_q \Phi_d \Omega_r]^T - [x1 \times x2 \times x3]^T \) has been introduced. \( \Omega_r \) is assumed to be constant during a sampling period: \( \Delta t = 0 \). The corresponding reduced-order extended state space equation becomes:

\[
\begin{align*}
\dot{x}_1 &= x_3 L_n I_n - x_1 x_1 + \omega_b x_2 \\
\dot{x}_2 &= x_3 L_n I_n - \omega_b x_1 - x_2 x_2 \\
\dot{x}_3 &= 0
\end{align*}
\]  

(7)

The fact that the state vector only consists of the rotor flux and resistance offers an advantage namely the reduction of the computational volume and complexity. Thus the rotor flux and resistance can be easily and rapidly estimated.

Denote \( \dot{x}_1 \) and \( \dot{x}_2 \) and \( \dot{x}_3 \) the estimates of the fluxes and rotor time constant. The proposed RESMO is a copy of the model (7) by adding corrector gains with switching terms:

\[
\begin{align*}
\dot{s}_1 &= \dot{s}_2 L_n I_n - s_1 \dot{s}_1 + \omega_b \dot{s}_2 + \Gamma_1 L_n \\
\dot{s}_2 &= \dot{s}_3 L_n I_n - \omega_b \dot{s}_1 - \dot{s}_2 \dot{s}_2 + \Gamma_2 I_n \\
\dot{s}_3 &= \Gamma_3 I_n
\end{align*}
\]  

(8)

where, \( \Gamma_1 \) and \( \Gamma_2 \) are the observer gains defined as (6). The switching \( I_n \) is in the form,

\[
I_n = \begin{bmatrix} \text{sign}(s_1) \\ \text{sign}(s_2) \end{bmatrix}
\]

with \( S = \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} = M \dot{Z} \) and \( M = \begin{bmatrix} \beta \sigma_1 & \beta \omega_1 \\ -\beta \omega_1 & \beta \sigma_1 \end{bmatrix} \)  

(9)

where, \( \dot{Z} \) is a function depending on the measures of stator currents, voltages and speed.

In order to determine the observer gain matrix \( \Gamma_2 \), it can be supposed that the observation errors of the fluxes converge to zero. The estimation error dynamics of the fluxes \( \dot{s}_1 = x_1 - \dot{x}_1 = 0 \) (i = 1, 2) are then given by:

\[
\begin{align*}
0 &= -s_2 \dot{s}_1 + s_1 \dot{s}_2 + \omega_b s_2 + L_n I_n s_3 - \Gamma_1 L_n \\
0 &= -\omega_b s_1 - s_2 \dot{s}_2 + s_1 \dot{s}_3 + L_n I_n s_3 - \Gamma_2 I_n \end{align*}
\]

By replacing the expressions of \( \Gamma_1 \) and \( \Gamma_2 \), we obtain:

\[
\begin{bmatrix} \dot{s}_1 \\ \dot{s}_2 \end{bmatrix} = \frac{1}{q} \begin{bmatrix} L_n I_n - \dot{s}_1 \\ L_n I_n - \dot{s}_2 \end{bmatrix} \dot{s}_3
\]

(10)

The estimation error dynamics of the rotor time constant is given by:

\[
\begin{align*}
\dot{s}_3 = -\Gamma_1 I_n = -\Gamma_1 \Delta_t \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} = -\Gamma_1 \Delta t \begin{bmatrix} L_n I_n \dot{s}_1 \\ L_n I_n - \dot{s}_3 \end{bmatrix} \dot{s}_3
\end{align*}
\]

We can see that this error dynamics is locally and exponentially stable by chosen:
\[ \Gamma_s = m q \left[ \begin{array}{c} \mathbf{I}_s \mathbf{I}_a - \mathbf{S}_s \\ \mathbf{I}_s \mathbf{I}_a - \mathbf{S}_s \end{array} \right] \Delta \quad \text{with} \quad m > 0 \] (11)

The parameter \( m \) is adjusted with respect to rotor time constant estimation.

In order to implement the RESMO algorithm in a DSP for real-time applications, the corresponding three-dimension state space equation defined in Eq. 7 must be discretized using Euler approximation (2nd order) proposed by Lewis (1992). Thus the new discrete-time varying model represented by a function depending on the stator current is given by:

\[
\begin{cases}
\dot{x}(k+1) = x(k) + T_s J_s(x(k),v(k)) + \frac{T_s^2}{2!} J_s(x(k),v(k)) \\
y(k) = x(k)
\end{cases}
\] (12)

where, \( x(k) = [\Phi_s(k) \Phi_p(k) \sigma_s(k)]^T \), \( v(k) = [I_s(k) I_p(k)]^T \)

\[
J_s(x(k),v(k)) = \begin{bmatrix}
-\sigma_s(k)\Phi_s(k) + \sigma_s(k)\Phi_p(k) + \sigma_s(k)L_s I_s(k) \\
-\sigma_p(k)\Phi_p(k) - \sigma_p(k)\Phi_s(k) + \sigma_p(k)L_p I_p(k) \\
0
\end{bmatrix}
\]

\[
J_s(x(k),v(k)) = \begin{bmatrix}
(\sigma_s^2(k) - \sigma_p^2(k))\Phi_s(k) - 2\sigma_s(k)\sigma_p(k)\Phi_s(k) \\
-\sigma_s^2(k)L_s I_s(k) + \sigma_s(k)L_p I_p(k) \\
0
\end{bmatrix}
\]

\[
J_s(x(k),v(k)) = \begin{bmatrix}
2\sigma_p(k)\sigma_s(k)\Phi_p(k) + (\sigma_p^2(k) - \sigma_s^2(k))\Phi_p(k) \\
-\sigma_s(k)L_p \sigma_p(k) I_p(k) - \sigma_p(k)L_s \sigma_s(k) I_s(k) \\
0
\end{bmatrix}
\]

where, \( k \) means the \( k \)th sampling time, i.e., \( t = k T \) with \( T \), the adequate sampling period chosen without failing the stability and the accuracy of the discrete-time model.

The proposed RDSMO can be defined by the following equation:

\[
\dot{x}(k+1) = \hat{x}(k) + T_s J_s(\hat{x}(k),v(k)) + \frac{T_s^2}{2!} J_s(\hat{x}(k),v(k)) + G(k) I_s(k)
\] (13)

where, the prediction vector is:

\[
\hat{x}(k+1) = \hat{x}(k) + T_s J_s(\hat{x}(k),v(k)) + \frac{T_s^2}{2!} J_s(\hat{x}(k),v(k))
\]

with \( \hat{x}(k) = [\hat{\Phi}_s(k) \hat{\Phi}_p(k) \hat{\sigma}_s(k)]^T \)

The switching vector \( I_s(k) \), deduced from the continuous case given by Eq. 9, can be written as:

\[
I_s(k) = \begin{bmatrix}
\text{sign}(s_1(k)) \\
\text{sign}(s_2(k))
\end{bmatrix} \quad \text{with} \quad S = \begin{bmatrix}
s_1(k) \\
s_2(k)
\end{bmatrix} = T_s M(k) \hat{Z}(k+1)
\] (14)

Where:

\[
M(k) = \begin{bmatrix}
-\hat{\sigma}_s(k) - \frac{T_s}{2} (\hat{\sigma}_s^2(k) - \sigma_s^2(k)) \\
\sigma_p(k) - T_s \hat{\sigma}_s(k) \sigma_p(k)
\end{bmatrix}
\]

\[
\hat{Z}(k+1) = \begin{bmatrix}
\hat{z}_s(k+1) \\
\hat{z}_p(k+1)
\end{bmatrix} = \begin{bmatrix}
z_s(k+1) - \hat{z}_s(k+1) \\
z_p(k+1) - \hat{z}_p(k+1)
\end{bmatrix}
\]
Setting $z_a(k+1) = \Phi_a(k+1) - \Phi_a(k) - T_\sigma \omega(k) \Phi_a(k)$ and $z_q(k+1) = \Phi_q(k+1) - \Phi_q(k) + T_\sigma \omega(k) \Phi_q(k)$, from the electrical Eq. 1a, an approximate discrete-time relation of the fluxes is given by:

$$
\begin{align*}
\begin{bmatrix}
z_a(k+1) \\
z_q(k+1)
\end{bmatrix} &= \begin{bmatrix}
\frac{L_a}{T_a} [V_a(k) - R_a I_a(k)] - \frac{\sigma L_a}{T_a} [I_a(k+1) - I_a(k) - T_\sigma \omega(k) I_p(k)] \\
\frac{L_q}{T_q} [V_q(k) - R_q I_q(k)] - \frac{\sigma L_q}{T_q} [I_q(k+1) - I_q(k) + T_\sigma \omega(k) I_a(k)]
\end{bmatrix}
\end{align*}
$$

and

$$
\begin{align*}
\begin{bmatrix}
z_a(k+1) \\
z_q(k+1)
\end{bmatrix} &= \begin{bmatrix}
\Phi_a(k+1) - \Phi_a(k) - T_\sigma \omega(k) \Phi_a(k) \\
\Phi_q(k+1) - \Phi_q(k) + T_\sigma \omega(k) \Phi_q(k)
\end{bmatrix}
\end{align*}
$$

(15)

The proposed gain matrix representation $G(k)$, deduced from the continuous case given by Eq. 6 and 11, can be defined as follows (discrete-time approach):

$$
G(k) = T_k \begin{bmatrix}
\Gamma_1(k) \\
\Gamma_2(k) \\
\Gamma_3(k)
\end{bmatrix} = \begin{bmatrix}
q - T_\sigma \hat{\sigma}_\omega(k) & T_\sigma \omega(k) \\
-T_\sigma \omega(k) & q - T_\sigma \hat{\sigma}_\omega(k) \\
T_\sigma (mL_n I_a(k) - \Phi_a(k)) & T_\sigma (mL_n I_q(k) - \Phi_q(k))
\end{bmatrix} \Delta
$$

(16)

Finally, from the Eq. 16, it can be seen that there are three positive adjusting gains: $q$, $n$ and $m$ which play a critical role in the stability and the velocity of the observer convergence. These three adjusting parameters must be chosen so that the reduced observer satisfies robustness properties, global or local stability, good accuracy and considerable rapidity.

Once the fluxes are estimated, it is easy to deduce the estimated torque defined by:

$$
\dot{\hat{C}}_{\omega}(k) = p \frac{L_n}{T_n} (\Phi_a(k) I_p(k) - \Phi_q(k) I_a(k))
$$

(17)

RESULTS

In order to verify the feasibility of the proposed RDES, the simulation on SIMULINK from Mathwork has been carried out for a 5.5 kW induction motor controlled with a field oriented vector strategy (Fig. 2). The nominal parameters of the induction motor are given in the Table 1.

The RDES is implanted in a S function using C language. In order to evaluate its performances and effectiveness, the comparisons between the observed state variables and the simulated ones have been realized for several operating conditions with the presence of about 20% noise on the simulated currents. Thus, using a sampling period $T_e = 1$ ms, the simulations are obtained at first in the nominal case with the nominal parameters of the IM (Table 1) used to realize vector control orientation and then, in the second case, with 50% variation of the nominal rotor time constant ($\sigma = 1.5 \sigma_n$) in order to verify the rotor time constant tracking and flux estimation. Two RST controllers are placed in the current loops $I_a$ and $I_q$, in order to realize the regulation of the flux and torque current, respectively.

<table>
<thead>
<tr>
<th>Table 1: Nominal parameters of the induction motor</th>
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<tbody>
<tr>
<td>$P_n = 5.5$ kW</td>
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<tr>
<td>$f_n = 50$ Hz</td>
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<tr>
<td>$R_n = 1$ Q</td>
</tr>
<tr>
<td>$L_n = 0.0037$ H</td>
</tr>
</tbody>
</table>

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Figure 3 and 4 show the simulation results for a step variation of the currents reference ($I_{ref}^q$ and $I_{ref}^d$). In the case of rotor resistance variation where $R_r = 1.5 R_{r0}$ (Fig. 4), one can see perturbations on the flux orientation when the current $I_q$ varies in order to generate the required torque. However, these waveforms remain acceptable and show that in both nominal (Fig. 3d) and non-nominal cases (Fig. 4d), the observed values of fluxes, rotor time constant and torque converge very well to their simulated values.

Finally, the implementation in real-time of the proposed scheme is carried out on a testing bench.

Figure 5 shows the experimental set-up. It is composed of a 5.5 kW induction motor, a powder brake with load torque measurements, three LEM current sensors and a 2000 point incremental encoder. A PC-board and a Dspace1102 combined a TMS320C31/40 MHz are used to implement PWM function and control algorithms. Our proposed RDESMO has been implemented using Euler approximation. The PWM and the position measurement work at 10 kHz. Field-oriented control and RDESMO operate at 1 m/sec sampling period.
Fig. 4: Non Nominal case (R_e = 1.5*R_n) with the presence of noises

Fig. 5: Experimental configuration diagram

15 kW three phase static inverter is supplied by a voltage source which provides about 0-400 V with current limitation of about 6 A.

Two kinds of tests have been performed (without and with load torque) in order to compare the behaviour of the reduced observer algorithm with respect to parameter variation:

- Figure 6 shows the simulation and experimental results at a constant speed of 700 rpm, with a stator current I_s = 4.3 A without load torque (C_t = 0 N m)
- Figure 7 shows the results where the motor speed is regulated at 1000 rpm with a load torque C_t = 3 N m and a stator current I_s = 2.5 A

For each test, the comparative simulation and experimental results are presented. Better estimation performance yielded by the proposed reduced order sliding mode observer is obvious from the experimental results. Thus it can be seen that the experimental waves are quite similar to the simulation ones. The experimental observed fluxes (Fig. 6b, 7a) indicates the good orientation (the flux \( \Phi_i \) is constant and \( \Phi_e \) converges to zero) which is due to a favourable rotor time constant estimation (Fig. 7b). Here the rotor time constant effectively
Fig. 6: Results for regulating the motor speed to 700 rpm without load torque $C_l = 0\, \text{N m}$

Fig. 7: Results for regulating the motor speed to 1000 rpm with a load torque $C_l = 3\, \text{N m}$

drives with the overheating of the IM because its estimated value is superior to the nominal one ($\alpha_0 = 10.16\, \text{sec}^{-1}$). The experimental estimated torque (Fig. 6a, 7c) is in good agreement with the simulated value. The weak perturbations on the experimented fluxes or torque are probably tied to position noises and the inverter.

The agreement between the experimental dynamic performance and the simulated ones is demonstrated.

CONCLUSIONS

A new approach for robust flux estimation on high-performance IM drive, namely, in sensorless control, was presented in this study. It is based on a reduced order extended sliding mode observer algorithm and on an innovative methodology used in the state-space model discretization.

The proposed RDESMO was successfully implemented for a direct rotor flux oriented IM drive. Very important and practical aspects and new improvements were introduced that strongly reduce the execution time of this new algorithm and simplify the tuning of gain matrices. In fact, the execution time of the RDESMO algorithm is about half of the full order extended SMO (Asseu et al., 2008).
The interesting simulation and experimental results obtained on the induction motor show the good performance and robustness of the RDES MO algorithm with respect to the rotor resistance variations, noises and load. The main contribution of this work is that the well-known drawbacks of the full order extended SMO, like heavy computational effort for real-time applications, complexity and hard tuning of the algorithm are widely overcome using the proposed RDES MO.

NOMENCLATURE

\begin{align*}
C_{mr}, C_i & : \text{Electromagnetic and load torques (N m)} \\
I_{as}, I_{at} & : \text{Stationary frame (d, q)-axis stator currents (A)} \\
I_{q}, I_{tr}, I_{wu} & : \text{Stationary frame (d, q)-axis rotor currents and rotor magnetizing current (A)} \\
p, J, f & : p: \text{pole pair No.}, J: \text{inertia (kg m^2)}, f: \text{friction coefficient (Nm sec rad}^{-1}) \\
L_r, L_{sr}, L_{mu}, L_{cu} & : \text{Roter, stator, mutual and leakage inductances (H)} \\
R_s, r_f & : \text{Stator and rotor referred resistance (} \Omega \text{)} \\
T_p, T_r, T_s & : \text{Sampling period, rotor and stator time constant: } T_r = L_r / R_s, T_s = L_s / R_p (s) \\
V_{as}, V_{sy} & : \text{Stationary frame d- and q-axis stator voltage (V)} \\
\Phi_{ds}, \Phi_{qs}, \Phi_{at}, \Phi_{wt}, & \text{d-q components of rotor fluxes (} \Phi_{a}, \Phi_{q} \text{) and stator fluxes (} \Phi_{a}, \Phi_{q} \text{), (Wb)} \\
\omega_a, \omega_s, \omega_d & : \text{Stator, rotor and slip pulsation (or speed), (rad sec}^{-1})
\end{align*}

REFERENCES


