NOVEL SPEED SENSORLESS INDIRECT FIELD-ORIENTED CONTROL OF INDUCTION MOTOR USING PLL AND EKF

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Abstract: *This paper presents an improved method of rotor flux angle θ_s estimation for sensorless indirect oriented field control of induction motors based on a phase locked loop (PLL). A PLL is used to estimate the rotor flux angle taking account only the reference speed. An extended Kalman Filter (EKF) is used to estimate speed and rotor flux. Extensive simulation and experimental tests confirm the effectiveness of the proposed approach.*

Keywords*: Induction motor (IM), phase locked loop (PLL), indirect field Oriented Control (IFOC), kalman filter, sensorless control.*

NOMENCLATURE

$[V_{sd}, V_{sq}]$:	Stator voltage in synchronous reference.
$[i_{sd}, i_{sq}]$:	Stator current in synchronous reference.
$[\emptyset_{rd}, \emptyset_{rq}]$	Rotor fluxes in synchronous reference.
$[V_{s\alpha}, V_{s\beta}]$	Stator voltage in stationary reference.
$[i_{s\alpha}, i_{s\beta}]$	Stator current in stationary reference.
$[\emptyset_{r\alpha}, \emptyset_{r\beta}]$	Rotor fluxes in stationary reference.
ω	Rotor speed.
ω_s	Synchronous frequency.
p	Number of pole pairs.
R_s	Stator resistance.
R_r	Rotor resistance.
R	Total resistance brought back to stator.
M	Magnetizing inductance.
$[L_s, L_r]$	Stator and rotor total inductances.
T_r	Rotor time constant.
J	Moment of inertia.
σ	Leakage factor.
f	Damping coefficient.
T_e	Electromagnetic torque.
T_l	Load torque.

1. INTRODUCTION

The indirect field oriented control (IFOC) technique is very useful for implementing high performance induction motor drive systems [l-5]. In the IFOC technique the shaft speed, that is usually measured, and the slip speed, that is calculated based on the machine parameters, are added to define the angular frequency of the rotor flux vector. Then, the standard IFOC technique is essentially a feed forward scheme which has the drawbacks being dependent on the parameters that vary with the temperature and the level of magnetic excitation of the motor. Using adaptive schemes for compensating the Parameter changes was proposed in [5],[6],[7],[8],[9],[10], [11], [12], [13].

In [6], the stator frequency is computed based on the induction-machine equations, in [10] the stator frequency is estimated with a phase locked loop (PLL). The PLL synthesizes a rotating reference frame that always tries to instantaneously align with the machine's voltage vector. The stator frequency is a stable by-product of the PLL process. The idea of using a PLL for frequency estimation comes from uninterruptible power supply (UPS) systems, The PLL only uses the voltage vector and is likely to yield a better frequency estimate under real conditions (imperfections, offsets, etc.) than the method based on the machine's equations.

This article investigates a technique (PLL) that uses a direct estimation of the frequency of the rotor flux vector (stator frequency) to obtain an IFOC independent of the torque and the rotor resistance. The estimate of the angular frequency of the rotor flux vector is obtained from the measurable stator variables (currents) and the reference speed. For testing this technique one has used an extended Kalman Filter estimator for speed and rotor flux estimation of induction motors [14],[15],[16].

Simulations and experimental results show that, the PLL strategy associated to sensorless IFOC control provide good results**.**

2. Induction motor model

The equivalent two-phase model of the symmetrical IM, under assumptions of linear magnetic circuits and balanced operating conditions, is expressed in an arbitrary rotating reference frame d-q) as given by equation (1)

$$\begin{cases} \frac{di_{sd}}{dt} = -\frac{R}{\sigma L_s} i_{sd} + \omega_s i_{sq} + \frac{M}{\sigma L_s L_r}\left(\frac{1}{T_r}\emptyset_{rd} + \omega\emptyset_{rq}\right) + \frac{1}{\sigma L_s} V_{sd} \\ \frac{di_{sq}}{dt} = -\frac{R}{\sigma L_s} i_{sq} - \omega_s i_{sd} + \frac{M}{\sigma L_s L_r}\left(\frac{1}{T_r}\emptyset_{rq} - \omega\emptyset_{rd}\right) + \frac{1}{\sigma L_s} V_{sq} \\ \frac{d\emptyset_{rd}}{dt} = \frac{M}{T_r} i_{sd} - \frac{1}{T_r}\emptyset_{rd} + (\omega_s - \omega)\emptyset_{rq} \\ \frac{d\emptyset_{rq}}{dt} = \frac{M}{T_r} i_{sq} - \frac{1}{T_r}\emptyset_{rq} - (\omega_s - \omega)\emptyset_{rd} \\ T_e = \frac{p.M}{L_r}\left(i_{sq}\emptyset_{rd} - i_{sd}\emptyset_{rq}\right) \end{cases} \qquad (1)$$

$$\frac{d\omega}{dt} = \frac{1}{J}(T_e - T_l + f.\omega) \qquad (2)$$

$$\sigma = 1 - \frac{M^2}{L_s L_r} \quad ; \quad R = R_s + \frac{M^2}{L_r^2} R_r \quad ; \quad T_r = \frac{L_r}{R_r}$$

Equation(3) give a model of induction machine drive in stationary reference frame $\alpha - \beta$.

$$\begin{cases} \frac{di_{s\alpha}}{dt} = -\frac{R}{\sigma L_s} i_{s\alpha} + \frac{M}{\sigma L_s L_r}\left(\frac{1}{T_r}\emptyset_{r\alpha} + \omega\emptyset_{r\beta}\right) + \frac{1}{\sigma L_s} V_{s\alpha} \\ \frac{di_{s\beta}}{dt} = -\frac{R}{\sigma L_s} i_{s\beta} + \frac{M}{\sigma L_s L_r}\left(\frac{1}{T_r}\emptyset_{r\beta} - \omega\emptyset_{r\alpha}\right) + \frac{1}{\sigma L_s} V_{s\beta} \\ \frac{d\emptyset_{r\alpha}}{dt} = \frac{M}{T_r} i_{s\alpha} - \frac{1}{T_r}\emptyset_{r\alpha} \\ \frac{d\emptyset_{r\beta}}{dt} = \frac{M}{T_r} i_{s\beta} - \frac{1}{T_r}\emptyset_{r\beta} \\ T_e = \frac{p.M}{L_r}\left(i_{s\beta}\emptyset_{r\alpha} - i_{s\alpha}\emptyset_{r\beta}\right) \end{cases} \qquad (3)$$

3. Indirect field oriented control

The aim of this work is to design and test a new speed sensorless controller, which allows operating at low speed range and exhibit good dynamic and steady state performances. PLL strategy is used to compute stator frequency. As shown in Eq (2) the expression of the electromagnetic torque in the dynamic regime presents a coupling between stator current and rotor flux. The main objective of the vector control of induction motors is, as in 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 then aligned with the rotor flux space vector. Under this condition we get:

$$\phi_{rd} = \phi_r \text{ and } \phi_{rq} = 0 \qquad (4)$$

The rotor dynamics are given by the following equations:

$$\frac{d\emptyset_r}{dt} = \frac{M}{T_r} i_{sd} - \frac{1}{T_r}\emptyset_r \qquad (5)$$

$$\frac{d\omega}{dt} = \frac{3p^2 M}{2JL_r}\left(\emptyset_r i_{sq}\right) - \frac{f}{J}\omega - \frac{p}{J} T_l \qquad (6)$$

$$T_e = \frac{3pM\emptyset_r}{2L_r} i_{sq} \qquad (7)$$

The standard orientation angle of rotor flux is given by:

$$\theta_s = \int \omega_s dt = \int \left(\omega + \frac{R_r M i_{sq}^*}{L_r \emptyset_r^*}\right) dt \qquad (8)$$

The rotor flux magnitude is related to the direct axis stator current by a first order differential equation so it can be controlled by controlling the direct axis stator current.

Under steady state operation rotor flux is constant, so (5) becomes

$$\emptyset_r = M i_{sd} \qquad (9)$$

The modified indirect vector control can be implemented using the following equations:

$$i_{sd}^* = \frac{\emptyset_r^*}{M}$$

(10)

$$i_{sq}^* = \frac{2}{3p}\frac{L_r}{M}\frac{T_e^*}{\emptyset_r^*} \qquad (11)$$

The orientation angle of rotor flux is estimated by PLL system.

The principal scheme of the modified indirect vector control is shown in figure (1) in which the function blocks F_1 and F_2 are presented by the equations (10) and (11) respectively.

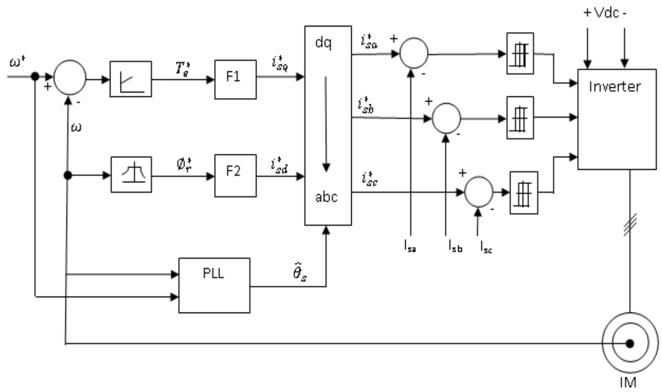

Fig. 1. Modified Indirect Field Oriented Control

In classical indirect FOC the rotor flux angle is generated in feed forward manner. Since this method relies on knowledge of the machine parameters such as M and $\frac{L_r}{R_r}$, these parameters could be affected by significant uncertainties and could vary during motor operations, thermal drift of stator and rotor resistance. The speed control in the indirect FOC in Figure (1) is achieved using a Proportional-Integral (PI) regulator.

3.1 Rotor Speed Regulation

The use of a classical PI controller makes appear in the closed loop transfer function a zero, which can influence the transient of the speed. Therefore, it is more convenient to use the so-called IP controller which has some advantages as a tiny overshoot in its step tracking response, good regulation characteristics compared to the proportional plus integral (PI) controller and a zero steady-state error, figure(1) shows the bloc diagram of PI speed controller. The speed filter accepts the speed signal as input and produces a modified speed signal for comparison to the speed reference signal ω^*. This completes the development of all the subsystems of the vector-controlled induction motor drive with constant rotor flux linkages. The design of PI speed controller is discussed below. The block diagram of speed control loop is shown in Figure(2).

$$T_e = K_T i_{sq}^* \tag{12}$$

$$i_{sq}^* = \frac{2}{3p}\frac{L_r}{M}\frac{T_e^*}{\emptyset_r^*} \tag{13}$$

$$i_{sd}^* = \frac{\emptyset_r^*}{M} \tag{14}$$

Where $K_T = \left(\frac{3pM^2}{2L_r}\right) i_{sd}^*$

Closed-loop transfer function with respect input is

$$\frac{\omega}{\omega^*} = \frac{K_i K_T}{Js^2+(f+K_pK_T)s+K_iK_T} = \frac{\omega_n^2}{s^2+2\delta\omega_n+\omega_n^2} \tag{15}$$

Where

$$\delta = \frac{f+K_pK_T}{2\sqrt{jK_iK_T}} \quad \text{and} \quad \omega_n = \sqrt{\frac{K_iK_T}{J}}$$

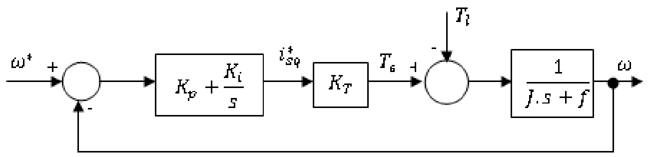

Fig.2. Block Diagram of Speed Control Loop

4. Design of the pll observer

The typical estimate of the stator field angle θ_s is computed using equation (8). The method uses the rotating reference frame torque current i_{sq} to compute the slip speed and depends on tshe inverse of the rotor time constant that varies with temperature..To overcome this problem a PLL can be used to estimate the angle (θ_s). The method uses a programmable low pass filter and a vector rotator to synthesize the stator fluxes. The stator frequency is estimated by a Phase Locked Loop (PLL). In essence, the flux estimation is still in the category of programmable LPF with vector rotator correction. The PLL synthesizes a rotating reference frame that always tries to instantaneously align with the voltage vector. The stator frequency is only a by-product of the PLL process [10]. A good estimation of the angle $(\hat{\theta}_s)$ depends on a good choice of VCO (Voltage Control Osciallator).

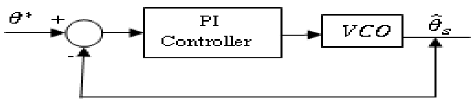

Fig.3. Proposed PLL for stator angle estimation in IM drives.

Transfer function $H(s)$, from input phase $\theta^*$to feedback phase $\hat{\theta}_s$, can be obtained as follows, the transfer function VCO is :

$$\left(\theta^* + \frac{Mi_{sq}^*}{s.T_r\emptyset_r^*}\right) \tag{16}$$

$$H(s) = \frac{(K_p.s+K_I).\left(\theta^*+\frac{Mi_{sq}^*}{s.T_r\emptyset_r^*}\right)}{1+(K_p.s+K_I)\left(\theta^*+\frac{Mi_{sq}^*}{s.T_r\emptyset_r^*}\right)} \tag{17}$$

The characteristic equation (18) is brought to analyzer the stability by Root locus technique.

$$1 + \left(K_p.s + K_I\right)\left(\theta^* + \frac{Mi_{sq}^*}{s.T_r\emptyset_r^*}\right) = 0 \tag{18}$$

5. Kalman filter observer

The model of the motor, after discretization is written as follows:

$$\begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \\ \phi_{r\alpha}(k+1) \\ \phi_{r\beta}(k+1) \end{bmatrix} = \begin{bmatrix} (1+a_1T_e) & 0 & a_2T_e & a_3T_e\omega(k) \\ 0 & (1+a_1T_e) & -a_3T_e\omega(k) & a_2T_e \\ a_4T_e & 0 & (1+a_5T_e) & -T_e\omega(k) \\ 0 & a_4T_e & T_e\omega & (1+a_5T_e) \end{bmatrix} . \begin{bmatrix} i_{s\alpha}(k) \\ i_{s\beta}(k) \\ \phi_{r\alpha}(k) \\ \phi_{r\beta}(k) \end{bmatrix} + \begin{bmatrix} bT_e & 0 \\ 0 & bT_e \\ 0 & 0 \\ 0 & 0 \end{bmatrix} . \begin{bmatrix} V_{s\alpha} \\ V_{s\beta} \end{bmatrix} \tag{19}$$

$$\begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} . \begin{bmatrix} i_{s\alpha}(k) & i_{s\beta}(k) & \phi_{r\alpha}(k) & \phi_{r\beta}(k) \end{bmatrix}^T$$

Where: $a_1 = \frac{1}{\sigma L_s} R$; $a_2 = \frac{M}{\sigma L_s L_r T_r}$; $a_3 = \frac{M}{\sigma L_s L_r}$; $a_4 = \frac{M}{T_r}$; $a_5 = \frac{1}{T_r}$; $a_6 = \frac{C_{cst}}{J}$; $a_7 = \frac{f}{J}$; $a_8 = \frac{pM}{JL_r}$

$b = \frac{1}{\sigma L_s}$.

Kalman Filter is a recursive mean squared estimator. It is capable of producing optimal estimates of system

states that are not measured. The elements of the covariance matrices Q
and R serve as design parameters for convergence of the system. The Kalman Filter approach assumes that the deterministic model of the motor is disturbed by centred white noise viz. the state noise and measurement noise. The discrete time-varying nonlinear model of the induction motor is of the following form [14],[15], [16]:

$$\begin{cases} X(k+1) = f(X(k),U(k),k) + b_{rs}(k) \\ Y(k+1) = C_d X(k) + b_{rm}(k) \end{cases} \tag{20}$$

where $f(X(k),U(k),k) = A_d X(k) + B_d U(k)$; and $b_{rs}(k)$ and $b_{rm}(k)$ are the system and measurement noises respectively.
These noises are supposed to be white with zero mean and characterized by:

$$\begin{gathered} E[b_{rs}(k)] = 0; E\{b_{rs}(k), b_{rs}(k)^T\} = Q(k) > 0 \\ E[b_{rm}(k)] = 0; E\{b_{rm}(k), b_{rm}(k)^T\} = R(k) > 0 \end{gathered} \tag{21}$$

The process covariance matrix *Q(k)* and measurement covariance matrix *R(k)* are symmetric and semi-definite. The EKF algorithm is as follows:
Step 1: Prediction

$$\hat{X}(k+1/k) = A(k).\hat{X}(k/k) + B(k)U(k) \tag{22}$$

$$P(k+1/k) = A(k).P(k/k).A(k)^T + Q(k) \tag{23}$$

where $\hat{X}(k)$ is the state estimate, *P(k)* is the estimation error covariance matrix.
Step 2: Updating

$$\hat{X}(k+1/k+1) = \hat{X}(k+1/k) + K_f(k+1).(Y(k+1) - \hat{Y}(k+1/k)) \tag{24}$$

$$K_f(k+1) = P(k+1/k).C^T (C.P(k+1/k).C^T + R(k+1))^{-1} \tag{25}$$

$$P(k+1/k+1) = (I - K_f(k+1).C).P(k+1/k) \tag{26}$$

where K_f is the Kahnan gain matrix, *(k+l/k)* denotes prediction at time *(k+1)* based on data up to and including *k*.

6. Simulation results

The proposed speed sensorless control algorithm has been tested by means of simulations and experiments using a 1.5 kW induction motor whose rated data are reported in Table I.

The proposed speed estimation algorithm is validated by simulation using Simulink. The block diagram of the sensorless indirect field oriented control of induction motor drive incorporating the speed estimator combined to a PLL approach is shown in Fig. 7 The accuracy of the estimation algorithm and response of the sensorless indirect field oriented control of induction motor drive is verified under fully loaded condition at various operating speeds. The results of simulation are shown in Fig. 8. These results are obtained by the simulation under the condition that the load torque is 1.5 (Nm). The motor speed (ω) , estimated speed $(\hat{\omega})$ and reference speed (ω^*) are shown in Fig.8.a. The estimated speed always followed the reference. The step change of the speed reference does not affect the performance of the system.
Fig.8-a show that the speed tracking errors are compensated by the proposed approach based on a PLL scheme combined to sensorless control. As a result estimation of the stator angle is achieved by a PLL mechanism. Good speed estimation accuracy was obtained under both dynamic and steady state conditions under various operating conditions.

Table.1

R_s : 5.72 Ω	STATOR RESISTANCE
R_r : 4.2 Ω	ROTOR RÉSISTANCE ROTOR
L_s : 0.462 H	STATOR INDUCTANCE
L_r : 0.462 H	ROTOR INDUCTANCE
M : 0.4402 H	MUTUAL INDUCTANCE
J : 0.0049 Kgm^2	MOMENT OF INERTIA
P : 2	NUMBER OF PAIR OF POLE
f : 0.003 Nm.s/rad	DAMPING COEFFICIENT
Power	1.5 KW

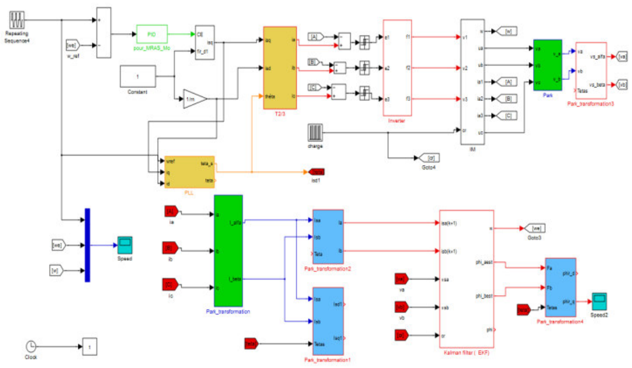

Fig.7. Simulation Block Diagram.

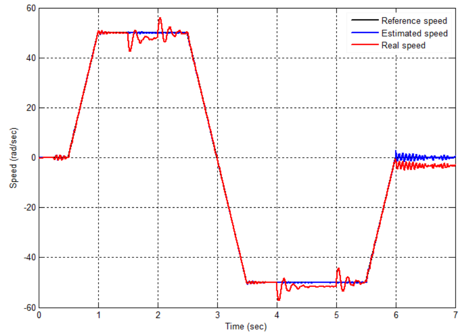

Fig.8-a . Real and estimated rotor speed.

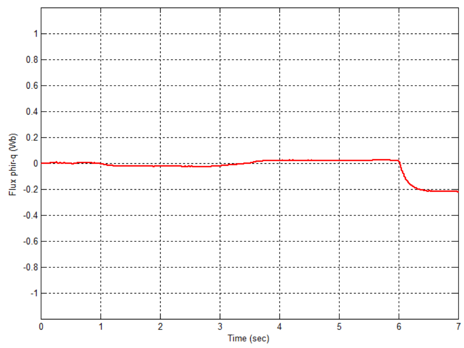

Fig.8-e . Direct flux $\widehat{\emptyset}_{rq}$.

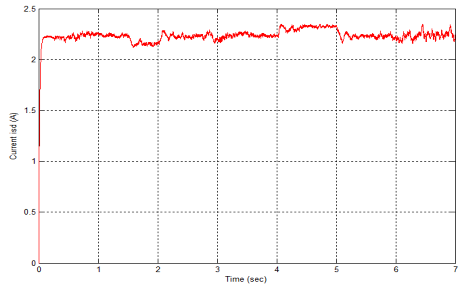

Fig.8-b . Direct current isd.

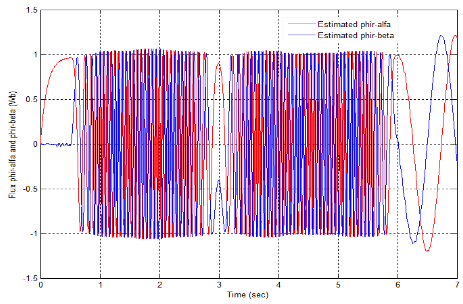

Fig. 8-f . Estimated rotor fluxes $\widehat{\emptyset}_{r\alpha}\widehat{\emptyset}_{r\beta}$.

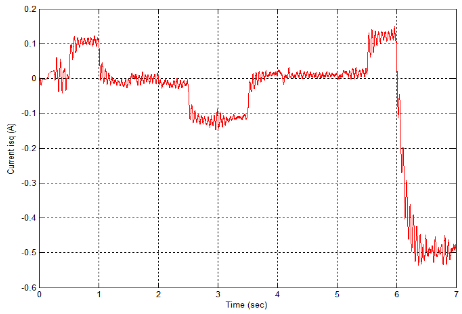

Fig. 8.c . Quadratic current isq.

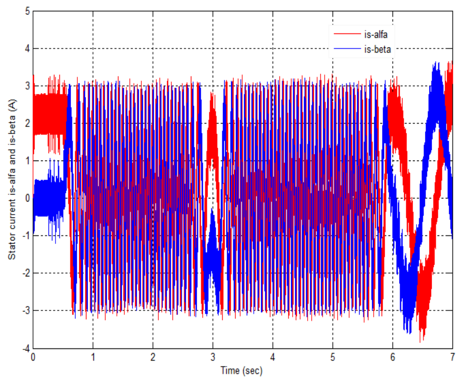

Fig. 8.g . Stator currents I_{sa} and I_{sb}.

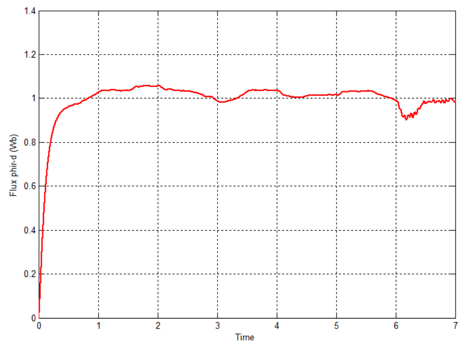

Fig.8.d . Direct flux $\widehat{\emptyset}_{rd}$.

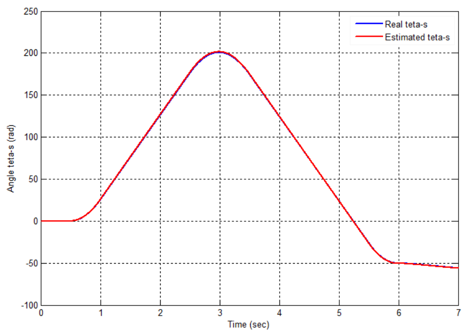

Fig. 8.h. Stator angle (stator frequency) estimation $\widehat{\theta}_s$.

7. Experimental results

The experimental system configuration is shown in Fig. 9 and its parameters are given in Table 1. The machine (1.5 Kw) is controlled by a DSP (rti1104, the laboratory (LTI,EA 3899) IUT of Soissons (french)) associated to a coprocessor (ADMC201) dedicated to the control of IM. The sampling frequency is fixed at 10 kHz and the controller receives the stator currents measurements through two 8- bit A/D converters. Then, using the PWM technique, the reference voltages are sent to the machine via the voltage-source inverter whose switching frequency is fixed at 10 kHz.
Figure 10a depicts the experimental motor speed and the estimated speed, while Figs. 10b and 10c show the real direct and quadratic current for a trapezoidal speed profile with a steady state at 50 rpm and a speed inversion. It can be noticed that the tracking capacities are good. Zero steady-state speed and rotor flux (fig 10e-f) estimation and tracking errors are achieved. Fig 10k depicts the estimated stator angle, from this result it can be noted good similarity between simulation (Fig8h) and experimental result.

Fig.9. The laboratory Motor-drive system.

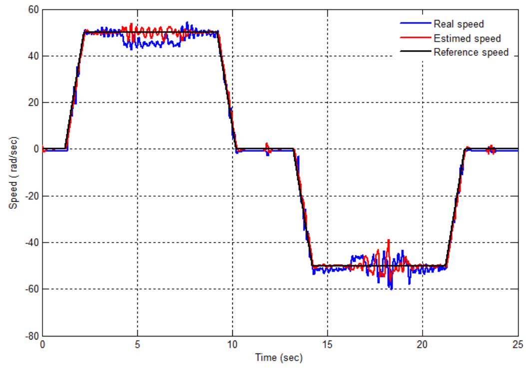

Fig.10.a Reference, estimated and real speed ($\pm 50\ rad/sec$).

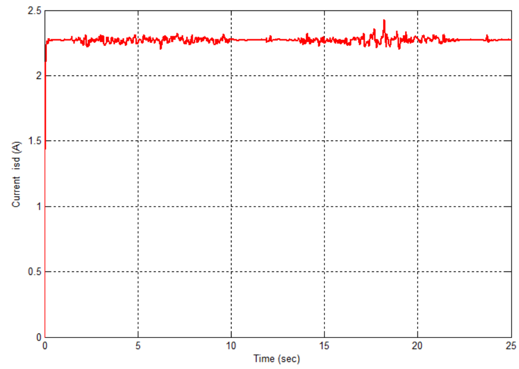

Fig.10.b. Direct current I_{sd}

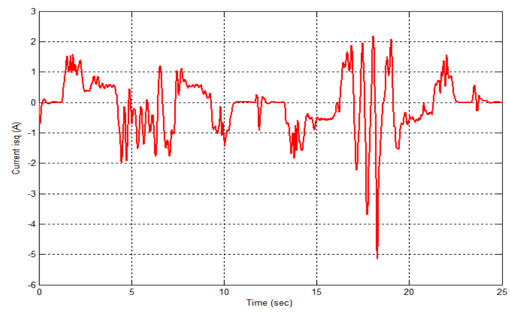

Fig.10.c. Quatratic current i_{sq}.

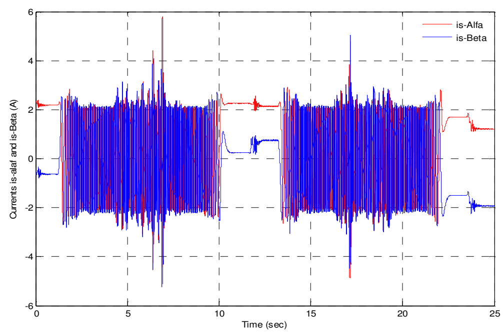

Fig.10.d. Stator currents I_{sa} and I_{sb}

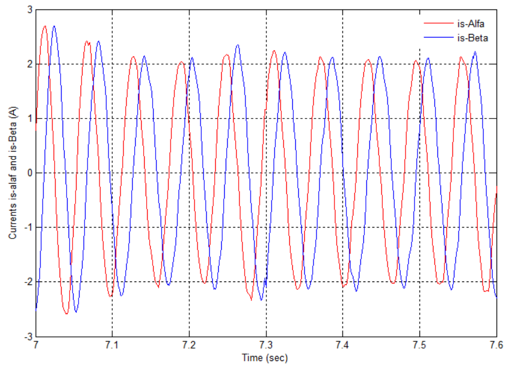

Fig.10.e. Stator currents I_{sa} and I_{sb} (Zoom).

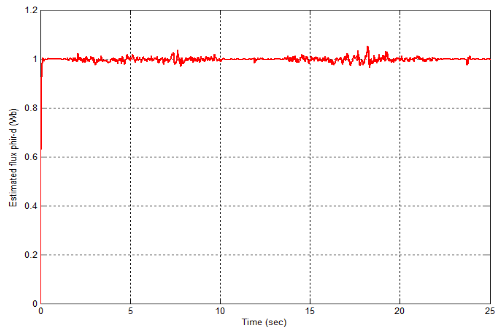

Fig.10.f. Direct estimated flux $\widehat{\emptyset}_{rd}$.

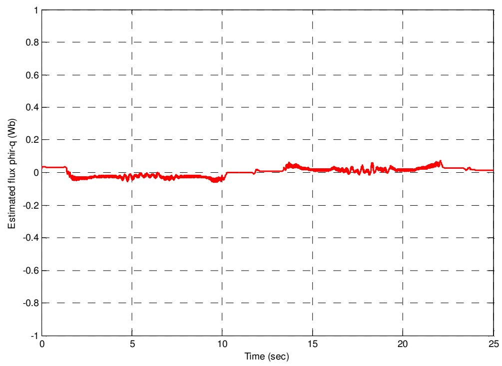

Fig.10.g. Quadratic current $\widehat{\emptyset}_{rq}$..

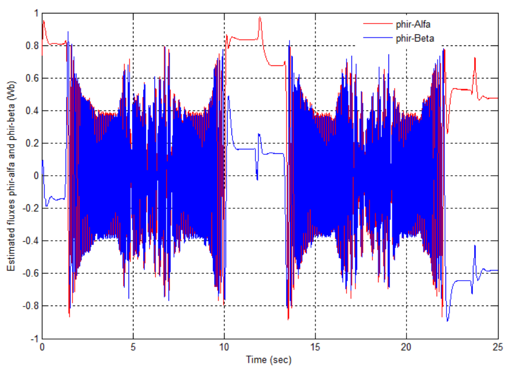

Fig 10.h. Rotor estimated fluxes $\widehat{\emptyset}_{r\alpha}$ *and* $\widehat{\emptyset}_{r\beta}$.

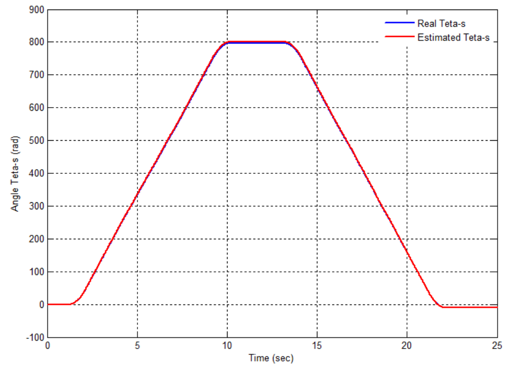

Fig.10.k. Stator angle (stator frequency) estimation $\hat{\theta}_s$.

8. Conclusion

In this paper, a speed sensorless controller for IM, which guarantees tracking of speed– flux references together with field orientation, has been presented. A extended Kalman Filter estimator for sensorless control of induction machines associated with an on-line estimation of the stator angle based on PLL scheme has been simulated and implemented successfully. Both simulation and experimental results have shown the performance of the method. Experiments and simulations results show that the proposed scheme with the stator angle estimation is suitable for high-performance sensorless controlled IM drives.

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