

A MMRAC CONTROLLER APPLIED TO ENCODERLESS SPEED CONTROL INDUCTION MOTOR DRIVES

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Abstract –This paper proposes a Modified Model Reference Adaptive Controller (MMRAC) for a speed sensorless induction motor drive. MMRAC is obtained by modifying the control law U of a Robust Model Reference Adaptive Controller (RMRAC). The MMRAC is used to assure performance in a wide speed range, including low- and zero-speed conditions. Experimental results are given to show the effectiveness of the proposed controller.

Keywords – Speed Sensorless, Induction Motor Drive, Recursive Least Square, RMRAC.

NOMENCLATURE

Plant variables

n_p	poles pair
V_{ds}	Direct stator voltage.
V_{qs}	Quadrature stator voltage.
I_{ds}	Direct stator current.
I_{qs}	Quadrature stator current.
I_{dr}	Direct rotor current.
I_{qr}	Quadrature rotor current.
ω	Synchronous reference frame.
ω_R	Rotor angular speed.
ω_{RM}	Model reference output.
L_S, L_R, L_M	Stator, rotor and mutual inductances.
R_S, R_R	Stator and rotor resistance.
B, J	Moment of inertia and damping coefficient.
$\bar{\sigma} = 1 - L_M^2 / (L_S L_R)$	

Speed estimator

Y, C	Prediction vector and regression matrix.
e	Identification error.
K	Gain vector.
P_R	Covariance matrix.
$P_1 = (R_S L_R + R_R L_S) / (L_S L_R \bar{\sigma})$	

Speed Controller MMRAC

U	Control law output.
w_1, w_2	Auxiliary filters.
$\theta_1, \theta_2, \theta_3, \theta_4$	Control law parameters.
P	Covariance matrix.
Ref	Speed reference.
m	Normalization signal.
σ	Sigma modification of control law parameters.
σ_R	Sigma modification of control law feedback.

I. INTRODUCTION

For a decade, induction motor (IM) drive-based electrical actuators have been under investigation as potential replacement for conventional hydraulic and pneumatic actuators in aircraft. Advantages of electric actuator include lower weight and size, reduced maintenance and operating costs, improved safety due to the elimination of hazardous fluids and high pressure hydraulic and pneumatic actuators, and increased efficiency.

Recently, research efforts have been devoted to eliminate speed sensor coupled to the shaft of the motor, presented in conventional closed loop servo systems. Motivations for substitution of speed sensor in estimation techniques are cost (usually a speed sensor is an expensive component) and reliability (this sensor is delicate and its signal can be interfered by electromagnetic sources).

In recent years, several encoderless vector control schemes have been proposed: algorithms using Kalman-filter [3]-[4], model reference adaptive systems [5]-[6], direct control of torque and flux [7]-[8], and linear models [12]-[1]. In Reyes et al [12] and Minami et al [1] a recursive algorithm is proposed to estimate the rotor speed based on measurements of stator voltages and currents. This technique is designed by two linear regression models derived from the machine electrical equations. However, all these techniques are based on back EMF measurement, and fail at low and zero speed because induced voltage levels are too low to be correctly measured. Moreover no voltages are induced on the stator windings at zero frequency.

Sensorless techniques based on estimation airgap flux position by using the third harmonic component of the stator voltage have been developed to improve the performance of EMF based DFOC drives [13]-[14]. These methods are based on the detection of the ripple generate on the angular frequency of the rotor flux by the injection of a suitable high frequency stator current signal. The rotor speed is estimated as a function of the rotor flux angular frequency ripple, the injected signal and the stator currents. Consequently these methods are motor parameters independent and allow estimating rotor speed at low and zero stator frequency, but these techniques can produce torque ripple and saturation in the main path and around the rotor slots, generating an additional modulation which interferes in the rotor speed estimation.

$$\begin{bmatrix} \dot{I}_{ds} \\ \dot{I}_{qs} \\ \dot{I}_{dr} \\ \dot{I}_{qr} \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{\bar{\sigma}L_s} & \omega + \frac{n_p\omega_R L_M^2}{\bar{\sigma}L_s L_R} & \frac{R_R L_M}{\bar{\sigma}L_s L_R} & \frac{n_p\omega_R L_M}{\bar{\sigma}L_s} \\ -\omega - \frac{n_p\omega_R L_M^2}{\bar{\sigma}L_s L_R} & -\frac{R_s}{\bar{\sigma}L_s} & -\frac{n_p\omega_R L_M}{\bar{\sigma}L_s} & \frac{R_R L_M}{\bar{\sigma}L_s L_R} \\ \frac{R_s L_M}{\bar{\sigma}L_s L_R} & -\frac{n_p\omega_R L_M}{\bar{\sigma}L_R} & -\frac{R_R}{\bar{\sigma}L_R} & \omega - \frac{n_p\omega_R}{\bar{\sigma}} \\ \frac{n_p\omega_R L_M}{\bar{\sigma}L_R} & \frac{R_s L_M}{\bar{\sigma}L_s L_R} & -\omega + \frac{n_p\omega_R}{\bar{\sigma}} & -\frac{R_R}{\bar{\sigma}L_R} \end{bmatrix} \begin{bmatrix} I_{ds} \\ I_{qs} \\ I_{dr} \\ I_{qr} \end{bmatrix} + \begin{bmatrix} \frac{1}{\bar{\sigma}L_s} & 0 \\ 0 & \frac{1}{\bar{\sigma}L_s} \\ -\frac{L_M}{\bar{\sigma}L_s L_R} & 0 \\ 0 & -\frac{L_M}{\bar{\sigma}L_s L_R} \end{bmatrix} \begin{bmatrix} V_{ds} \\ V_{qs} \end{bmatrix} \quad (1)$$

$$T_E = n_p L_M (I_{dr} I_{qs} - I_{ds} I_{qr}) \quad (2)$$

$$J \frac{d}{dt} \omega_R + B \omega_R = T_E - T_L \quad (3)$$

$$\begin{bmatrix} \dot{I}_{ds} \\ \dot{I}_{qs} \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{\bar{\sigma}L_s} & \omega \\ -\omega & -\frac{R_s}{\bar{\sigma}L_s} \end{bmatrix} \begin{bmatrix} I_{ds} \\ I_{qs} \end{bmatrix} + \begin{bmatrix} \frac{1}{\bar{\sigma}L_s} & 0 \\ 0 & \frac{1}{\bar{\sigma}L_s} \end{bmatrix} \begin{bmatrix} V_{ds} \\ V_{qs} \end{bmatrix} \quad (4)$$

$$T_E = \frac{n_p L_M^2 I_{qs} I_{ds}^*}{L_R} \quad (5)$$

To assure performance in a wide speed range, including low and zero speed conditions, this paper proposes the use of an estimator associated with the RMRAC control law in a speed sensorless IM servo system. A modification is proposed in the RMRAC technique to obtain the Modified Model Reference Adaptive Controller (MMRAC). The resultant controller is capable of ensuring good performance in a wide speed range, including low and zero speed conditions. Experimental results are presented to demonstrate the dynamic performance of the proposed closed-loop system.

II. PLANT DESCRIPTION

The dq model of the three-phase IM, with electrical variables referred to an arbitrary dq rotating frame, can be described by (1), (2) and (3), where the motor is considered perfectly balanced and regardless the saturation phenomena. $\bar{\sigma} = 1 - L_M^2 / (L_s L_R)$, V_{ds} , V_{qs} are the stator voltages; R_s , R_R are the stator and rotor resistance. L_s , L_R , L_M are the stator, rotor and mutual inductances; I_{ds} , I_{qs} , I_{dr} , I_{qr} are the stator and rotor currents; ω , ω_R are the rotating speed of the electrical current and of the shaft; T_E , T_L are the electrical torque and the load torque; J , B are the moment of inertia and the damping coefficient (motor and load); n_p is the number of pole pairs of the motor.

Equation (2) represents the coupling between the electrical and mechanical parts, and the mechanical part is given by the equation (3).

Applying IFOC techniques in (1) and (2) results in equations (4) and (5), where (*) define a reference signal, [11]. The I_{ds} current is kept constant to ensure a constant level of machine magnetization. This assumption is necessary to obtain linearization in (5). The frequency of the voltage applied to the IM is the rotor flux frequency given by

$$\omega = n_p \omega_R + (R_R I_{qs} / L_R I_{ds}) \quad (6)$$

Note that the rotor speed is required to obtain the synchronous speed and to convert the measurements of the stator voltages and currents to dq reference frame. Here it is assumed that speed sensor can not be used. Therefore an estimation algorithm is presented in the next section to obtain ω_R .

III. SPEED ESTIMATION ALGORITHM

Consider the three phase induction motor defined in (1)-(3) referred to the stator fixed frame.

$$\begin{bmatrix} V_{ds}^s \\ V_{qs}^s \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} a_1 & 0 & pL_M & 0 \\ 0 & a_1 & 0 & pL_M \\ pL_M & \omega_R L_M & a_2 & \omega_R L_M \\ -\omega_R L_M & pL_M & -\omega_R L_M & a_2 \end{bmatrix} \begin{bmatrix} I_{ds}^s \\ I_{qs}^s \\ I_{dr}^s \\ I_{qr}^s \end{bmatrix} \quad (7)$$

where $(^s)$ defines stationary referential variables, $a_1 = R_s + pL_s$, $a_2 = R_R + pL_R$ and $p = d/dt$. Using equation (7), it is possible to reformulate the problem of estimating speed as a problem of estimating the parameters based on a linear regression model. Linear regression models are models that have the structure given by

$$Y = C \hat{\omega}_R \quad (8)$$

Considering the rotor speed as the only unknown parameter, from (7), the linear model is given by

$$Y = \begin{bmatrix} \ddot{I}_{ds}^s + P_1 \dot{I}_{ds}^s + a_3 I_{ds}^s - \frac{1}{\bar{\sigma}L_s} \dot{V}_{ds}^s - \frac{R_R}{L_s L_R \bar{\sigma}} V_{ds}^s \\ \ddot{I}_{qs}^s + P_1 \dot{I}_{qs}^s + a_3 I_{qs}^s - \frac{1}{\bar{\sigma}L_s} \dot{V}_{qs}^s - \frac{R_R}{L_s L_R \bar{\sigma}} V_{qs}^s \end{bmatrix} \quad (9)$$

$$C^T = \begin{bmatrix} -\dot{I}_{qs}^s - \frac{R_s}{L_s \bar{\sigma}} I_{qs}^s + \frac{1}{L_s \bar{\sigma}} V_{qs}^s \\ \dot{I}_{ds}^s + \frac{R_s}{L_s \bar{\sigma}} I_{ds}^s - \frac{1}{L_s \bar{\sigma}} V_{ds}^s \end{bmatrix}$$

$$\frac{V_{fds}^s}{V_{ds}^s} = \frac{V_{fq_s}^s}{V_{qs}^s} = \frac{I_{fds}^s}{I_{ds}^s} = \frac{I_{fq_s}^s}{I_{qs}^s} = G_f(s) = \frac{\omega_c^3}{(s + \omega_c)^3} \quad (10)$$

where $P_1 = (R_s L_R + R_R L_s) / (L_s L_R \bar{\sigma})$, $a_3 = \frac{R_s R_R}{L_s L_R \bar{\sigma}}$ and $\omega_c = 5\omega$, [12].

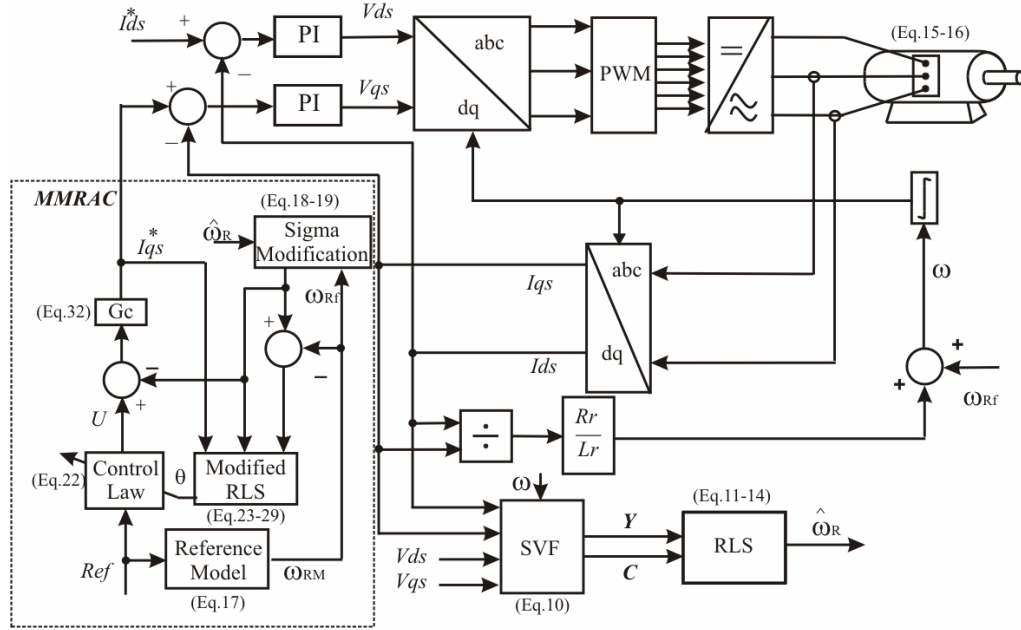


Fig. 1. Controller Structure

The regression model presented and other variations of it can be obtained from (7) with the electrical variables referred to the stator fixed frame as presented in [12]. It is assumed that the derivatives presented in (9) are measurable quantities. In the implementation these quantities are obtained by state variable filters (SVF) [16]. These filters are developed by discretization of transfer function given by (10) where the inputs V_{ds}^s , V_{qs}^s , I_{ds}^s and I_{qs}^s are used to obtain the filtered signals V_{fds}^s , V_{fqs}^s , I_{fds}^s and I_{fqs}^s .

To estimate $\hat{\omega}_R$ a recursive least squares algorithm can be used, [2]. The recursive equations are presented as follows

$$\hat{\omega}_R(t) = \hat{\omega}_R(t-1) + \mathbf{K}(t)\mathbf{e}(t) \quad (11)$$

$$\mathbf{e}(t) = \mathbf{Y}(t) - \mathbf{C}^T(t)\hat{\omega}_R(t-1) \quad (12)$$

$$\mathbf{K}(t) = \frac{\mathbf{P}_R(t-1)\mathbf{C}(t)}{1 + \mathbf{C}^T(t)\mathbf{P}_R(t-1)\mathbf{C}(t)} \quad (13)$$

$$\mathbf{P}_R(t) = 1 + \mathbf{K}(t)\mathbf{C}^T(t)\mathbf{P}_R(t-1) \quad (14)$$

By using this algorithm, it is possible to obtain the estimated speed $\hat{\omega}_R$ used in the control law presented in the next section.

IV. CONTROLLER STRUCTURE

The proposed sensorless speed control structure is shown in Figure 1. It uses current signals measured on the stator of the machine. A MMRAC control law presented in the next section is used to obtain the I_{qs}^* current calculated by the difference between the estimated speed $\hat{\omega}_R$ and the output of the reference model ω_{RM} . A RMRAC controller is employed to compensate parametric variations on the mechanical plant. Moreover, the use of a speed observer to obtain a signal to be controlled by the RMRAC controller results in a sensorless controller tolerant to unmodeled dynamics.

V. MMRAC CONTROLLER STRUCTURE

For a SISO plant (Single-Input Single-Output), a RMRAC controller can be designed as:-

$$\omega_R = G(s)U \quad (15)$$

$$G(s) = G_0(s)[1 + \mu\Delta_m(s)] + \mu\Delta_a(s) \quad (16)$$

where $G(s)$ is the plant transfer function, $G_0(s)$ is the modeled part of the plant and $\mu\Delta_m(s)$ and $\mu\Delta_a(s)$ are additive and multiplicative perturbations, respectively. The modeled part of the plant $G_0(s)$ is a strictly proper transfer function, expressed as $G_0(s) = k_p Z_0(s)/R_0(s)$, where $Z_0(s)$ and $R_0(s)$ are monic polynomials of degree m and n respectively. In addition, the following assumptions on $G_0(s)$ are made:

- S1) $R_0(s)$ is a monic polynomial of degree n and $Z_0(s)$ is monic Hurwitz polynomial of degree $m \leq (n-1)$;
- S2) The signal of k_p and the values of m and n are known.

Furthermore, for the unmodeled plant part is assumed that

- S3) $\Delta_a(s)$ is a strictly proper stable transfer function and $\Delta_m(s)$ is a stable transfer function;
- S4) A bound $p_0 > 0$ on the stability margin $p > 0$ for which the poles of $\Delta_a(s-p)$ and $\Delta_m(s-p)$ are stable is known.

The adaptive control objective can be described as follows. Given the reference model

$$G_M(s) = k_m Z_m(s)/R_m(s), \omega_{RM} = G_M(s)Ref \quad (17)$$

where $R_m(s)$ is a Hurwitz polynomial of degree $n^* = n - m$, where the Ref is a uniformly bounded reference, design an adaptive controller so that for some $\mu^* > 0$ and any $\mu \in [0, \mu^*]$, the resulting closed-loop plant is stable and the plant output $\hat{\omega}_R$ tracks the reference model output ω_{RM} as closely as possible, despite disturbances $\Delta_a(s)$ e $\Delta_m(s)$, satisfying S3 and S4.

As verified in [12], the speed estimator used has problems at very low speed. To overcome this problem, in this paper it is proposed the addition of a sigma modification in the RMRAC controller:

$$\hat{\omega}_{Rf} = \sigma_R \hat{\omega}_R + (1 - \sigma_R) \omega_{RM} \quad (18)$$

where

$$\sigma_R = \begin{cases} 0 & \text{if } \|\omega_{RM}\| < M_{R0} \\ \sigma_{R0} \left(\frac{\|\omega_{RM}\|}{M_{R0}} - 1 \right) & \text{if } M_{R0} \leq \|\omega_{RM}\| < 2M_{R0} \\ \sigma_{R0} & \text{if } \|\omega_{RM}\| \geq 2M_{R0} \end{cases} \quad (19)$$

The sigma modification defines when it is possible to use the $\hat{\omega}_R$ signal in the control law. When ω_{RM} is smaller than M_{R0} , the signal used in the control law feedback is ω_{RM} . On the other hand, when ω_{RM} is bigger than M_{R0} , the $\hat{\omega}_R$ signal is gradually employed in the control law feedback. When ω_{RM} is bigger than $2M_{R0}$, the $\hat{\omega}_R$ signal is appropriated to be used in the control law (22).

The input control law U and output $\hat{\omega}_{Rf}$ are used to generate $n-1$ dimensional auxiliary vectors so that

$$\dot{w}_1 = F w_1 + q U \quad (20)$$

$$\dot{w}_2 = F w_2 + q \hat{\omega}_{Rf} \quad (21)$$

where F is a stable matrix and the (F, q) is a controllable pair. The MMRAC signal U is given by

$$U = w_1 \theta_1^T + w_2 \theta_2^T + \hat{\omega}_{Rf} \theta_3 + Ref / \theta_4 \quad (22)$$

where θ_1^T , θ_2^T , θ_3 and θ_4 are control parameters. These parameters are obtained by a modified RLS presented in [10] and described as follows.

VI. MMRAC PARAMETER ESTIMATION ALGORITHM

In the proposed controller, a modified recursive least-squares (RLS) algorithm is used. This basic MMRAC algorithm structure is showed in Figure 2.

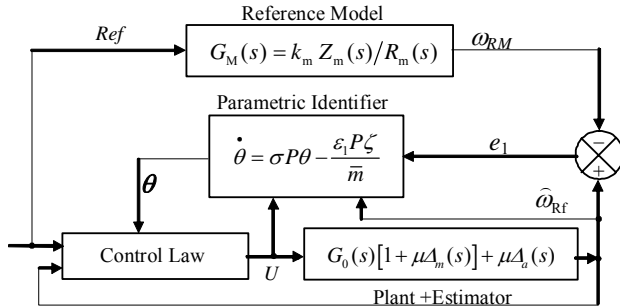


Fig. 2. Robust model reference adaptive control structure

The control law parameters are obtained by

$$\dot{\theta} = -\sigma P \theta - \frac{\varepsilon_1 P \zeta}{\bar{m}} \quad (23)$$

where $\theta = [\theta_1 \theta_2 \theta_3 \theta_4]^T$,

$$\dot{P} = \lambda \mu^{-2} P - \left(\frac{P \zeta \zeta^T P}{\bar{m}} + \bar{\mu}^2 \frac{P^2}{R^2} \right) \quad (24)$$

and $P = P^T$ is so that

$$0 < P(0) \leq \lambda R^2 I, \quad \mu^2 \leq k_\mu \bar{\mu}^2 \quad (25)$$

$$\bar{m} = 1 + \alpha_1 [m]^2, \quad \zeta = G_m I w \quad (26)$$

$$\dot{m} = \delta_0 m + \delta_1 (|U| + |\hat{\omega}_{Rf}| + 1), \quad m(0) > \frac{\delta_1}{\delta_0}, \quad \delta_1 \geq 1 \quad (27)$$

where α_1 , δ_0 , δ_1 , λ , $\bar{\mu}$ and R^2 are positive constants and δ_0 satisfies $\delta_0 + \delta_2 \leq \min[p_0, q_0]$. $q_0 \in \mathfrak{R}^+$ is such that the poles of $G_M(s-q_0)$ and the eigenvalues of $(F + q_0 I)$ are stable and $\delta_2 \geq 0$. $p_0 > 0$ is defined in S4 and σ in (23) can be expressed as

$$\sigma = \begin{cases} 0 & \text{if } \|\theta\| < M_0 \\ \sigma_0 \left(\frac{\|\theta\|}{M_0} - 1 \right) & \text{if } M_0 \leq \|\theta\| < 2M_0 \\ \sigma_0 & \text{if } \|\theta\| \geq 2M_0 \end{cases} \quad (28)$$

where $M_0 > \|\theta^*\|$ and $\sigma_0 > 2\bar{\mu}^2 / R^2 \in \mathfrak{R}^+$ are design parameters. As defined in [10], the modified error is given by

$$\varepsilon_1 = e_1 + \theta^T \zeta - G_M \theta^T w \quad (29)$$

VII. EXPERIMENTAL RESULTS

The MMRAC sensorless speed servo was implemented in a PC-based platform driving an induction motor. The motor is a Y-connected, one pole-pair, 0.9 Hp, 3400 rot/min, 380-V/2.7-A type. Motor parameters are obtained by no-load test, locked-rotor test and board data, and are presented in TABLE 1.

The design begins defining the reference model that it imposes the closed-loop dynamic of the system.

$$G_M(s) = \frac{8.2s + 0.01}{s^2 + 8.2s + 0.01} \quad (30)$$

Using the parameters presented in TABLE 1, the mechanical plant $G_0(s)$ is given by

$$G_0(s) = \frac{1/J}{s + B/J} = \frac{2702.7}{s + 2.702} \quad (31)$$

Note that there is a large difference between the dynamic of the model reference, and the dynamic of the plant, which may cause problems in the MMRAC controller. Moreover, it is necessary that $n > 1$ to enable the use of filters (20) and (21), [10]. To overcome these problems a pre-compensator $G_C(s)$ is used.

$$G_C(s) = \frac{0.009s + 0.01}{s} \quad (32)$$

The gains F and q used in (20) and (21) are equal to 10 and 1, respectively. The DC bus is limited in 170V/5A, as a consequence, the reference speed is 60 rad/s limited. The sampling time used is 555 μ s and the reference signal is kept at zero until 4s to ensure magnetization of the machine.

Before driving the proposed servo system, it is necessary to obtain experimentally the initial control law parameters $\theta(0)$ by using a tachometer to feedback the control law MMRAC. In this case, the estimated signal is replaced by the measured rotor speed ω_R and the constant σ_R is set equal to 1 during the test. The MMRAC parameters are given in TABLE 3 and $\theta(0) = [-5 \ 5 \ -5 \ -5]^T$, which norm is bigger than M_0 . Figure 3 presents the speed servo response using a speed sensor to a reference step. At the final of this test, the MMRAC control law parameters are obtained, Figure 4. Then $\theta(0) = [-2.63 \ 2.44 \ 1.82 \ -2.8]^T$ that will be used in the proposed encoderless servo system.

Figures 6-8 show the speed sensorless controller responses obtained using the reference presented in Figure 5. In a first test, the system is initially operating at 0 rad/s, when the speed reference is increased until 60 rad/s. After 10 sec, the speed reference is reduced to -30 rad/s. It can be seen that some spikes are presented in the speed estimated during low speed. These spikes cause some oscillations at d-axis and q-axis currents, as shown in Figure 7. Note that these oscillations do not degrade the control action, Figure 6.

Figures 10-12 show the proposed controller responses using the reference presented in Figure 9. The system is initially operating at 0 rad/s, when the speed reference is increased until 60 rad/s. After 10 sec, the speed reference is reduced to 0 rad/s, and finally it is increased to 30 rad/s. Figure 11 presents the control law parameters obtained and Figure 12 shows the direct and quadrature currents. The actual and estimated speed signals agree quite well for both steady state and transient condition and good accuracy is obtained for this condition, Figure 10.

Figures 13 and 14 present the results obtained in a comparison test between the proposed controller response and the RLS estimator proposed in [12]. In this case, several reference steps are used, including low and zero speed. The system is initially operating at 0 rad/s, when after 4 s, the speed reference is increased from 10 rad/s in each 8 s. Figure 13 presents the measured speed and model reference output. This response is obtained using the MMRAC controller. Note the good accuracy obtained for this condition. Figure 14 presents the estimated signal using the technique proposed in [12]. In speed reference above 10 rad/s the RLS estimator presents good response. However, in condition below 10 rad/s, a wrong estimated signal is presented, as can be verified in Figure 14. It disables the direct feedback of this signal. These results are obtained with no load.

For the load step test, a DC generator (the parameters of which are given in TABLE 2) is connected to the IM rotor. The system is initially operating at 0 rad/s, when the speed reference is increased to 60 rad/s. During the test, a 15 Ω resistance is connected on the generator terminals in 20 sec. Figure 15 shows the performance in an instantaneous load torque change and Figure 16 presents the quadrature current obtained.

TABLE 1
Motor Parameters

Inertia Moment	0.00037
Damping Coefficient	0.001
n_p	1
L_M	0.18 H
L_R	0.23 H
L_S	0.21 H
R_R	2.11 Ω
R_S	3.86 Ω
Nominal Current	3.50 A

TABLE 2
DC Generator Parameters

Power	500 W
Nominal Speed	1800 RPM
Max. Field Voltage	190 V
Used Field Voltage	100 V
Nominal Current	3.5 A

TABLE 3
MMRAC PARAMETERS

M_0	4
σ_0	1
F	4
Q	1
M_{R0}	5
σ_{R0}	1
$P_R(0)$	100
P(0)	300
δ_1	1
δ_0	1
δ_m	0

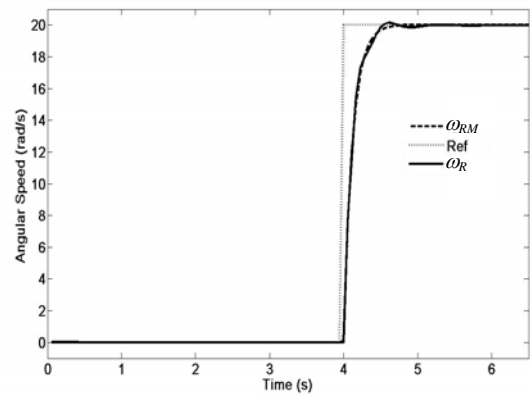


Fig. 3. Servo response using a speed sensor

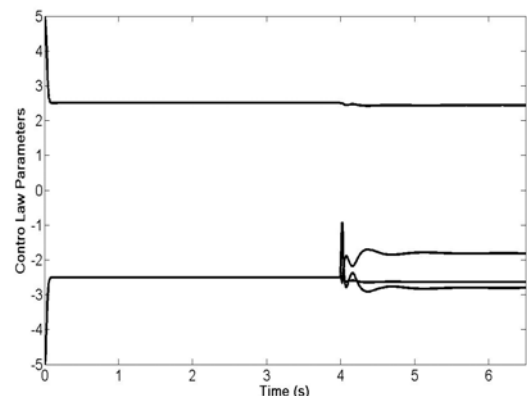


Fig. 4. Experimental parameters of the MMRAC control law

VIII. CONCLUSION

This paper described a modified model reference adaptive controller (MMRAC) applied to a speed sensorless servo system using three-phase induction motor. A speed estimator was incorporated in the control system to avoid the use of mechanical sensors. Experimental results demonstrated the effectiveness of the proposed control scheme, including the control at low speed and compensation of torque disturbances.

Moreover, the proposed scheme can be designed for a reduced order plant, without a priori knowledge of the exact model of the plant and the PWM inverter. In contrast with other RMRAC controllers, this scheme does not use measurements of the plant output signal to control it, but uses an observed signal $\hat{\omega}_R$.

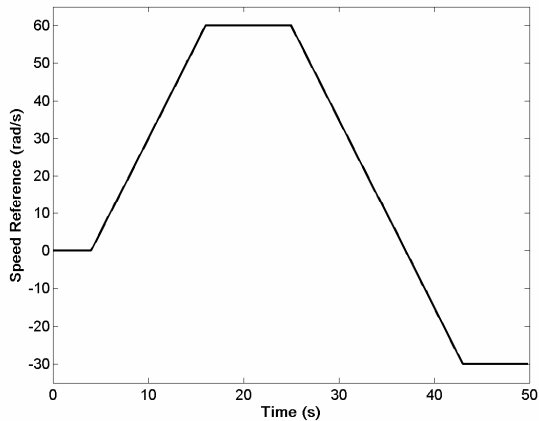


Fig. 5. Reference signal of the first test

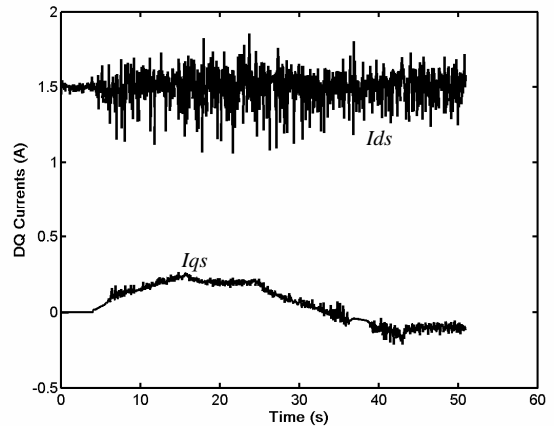


Fig. 8. Quadrature current obtained during the first test

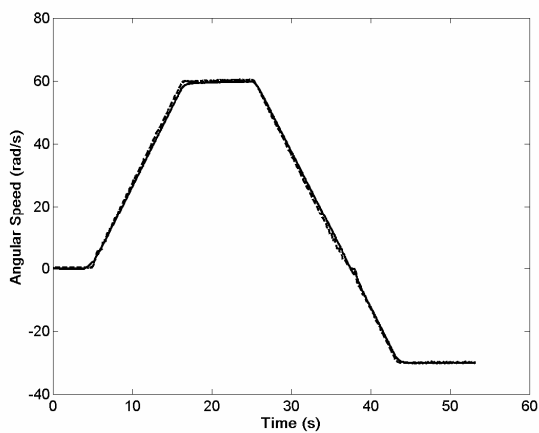


Fig. 6. Estimated speed (solid line) and measured speed (dashed line) in first test

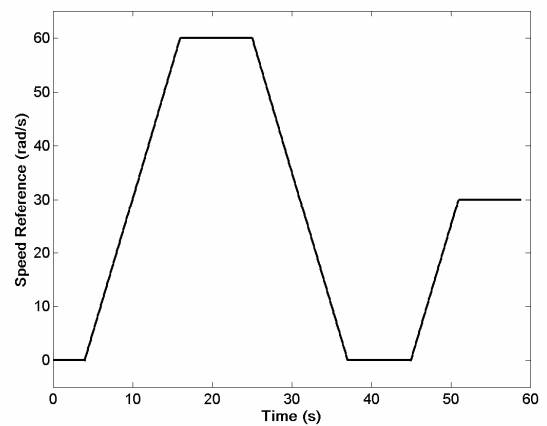


Fig. 9. Reference signal in the second test

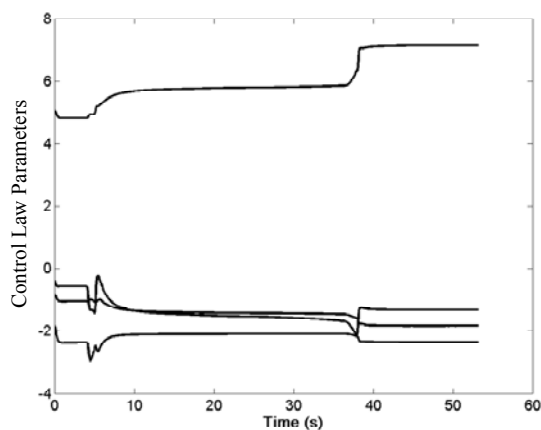


Fig. 7. Control law parameters during the first test

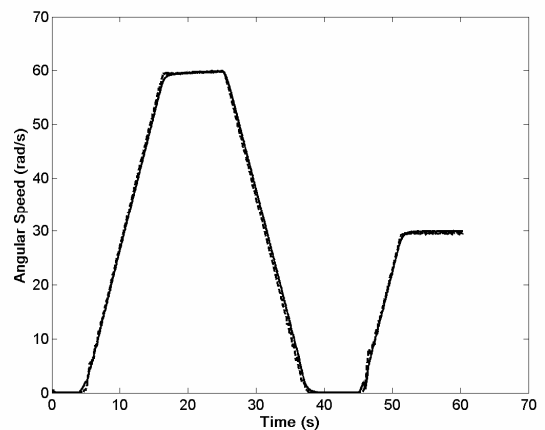


Fig. 10. Estimated speed (solid line) and measured speed (dashed line) in second test

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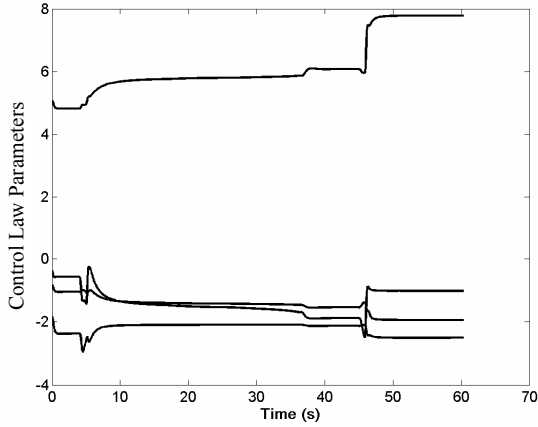


Fig. 11. Control law parameters during the second test

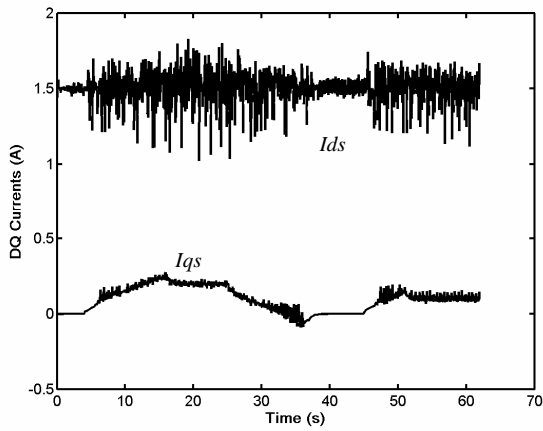


Fig. 12. Quadrature current obtained during the second test

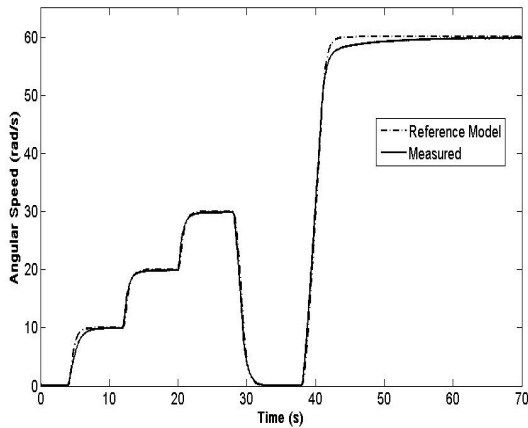


Fig. 13. Measure speed (solid line) and model reference output (dotted line) in a comparison test

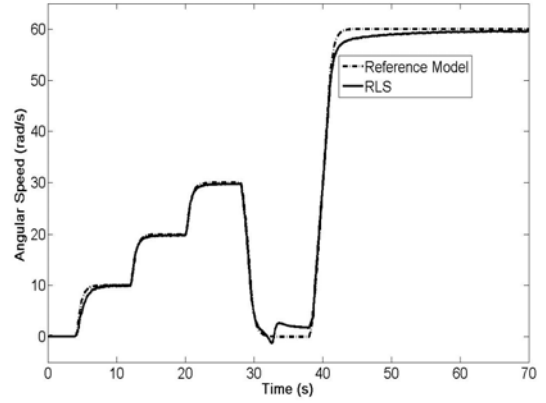


Fig. 14. Estimated speed (solid line) and model reference output (dotted line) in a comparison test

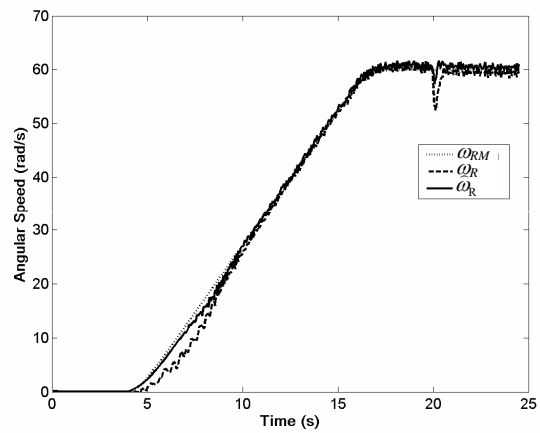


Fig. 15. Estimated speed (dashed line), measured speed (solid line) and model reference speed (dotted line) in torque disturbance test

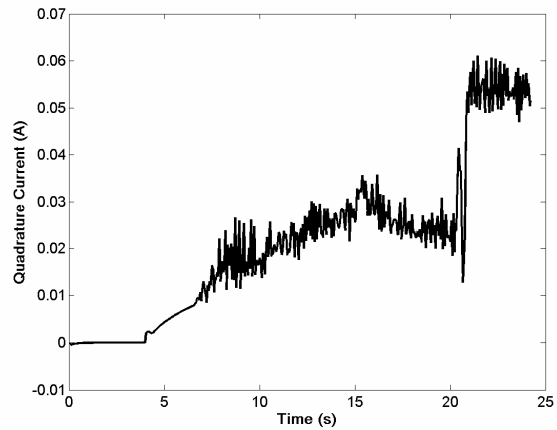


Fig. 16. Quadrature current obtained during the torque disturbance test

APPENDIX

Lemma 1: Let $C(t)$ given by (9) where V_{ds} , V_{qs} , I_{ds} , I_{qs} and them derivates are piecewise continuous function of time. Moreover, consider the linear error equation $e_L = (\hat{\omega}_R - \omega_R)C(t)$ (vide [2] eq. 2.4.3) with a RLS

algorithm presented in (8)-(14) and $\sigma_R = 1$. Defining a vector $C: \mathfrak{R}_+ \rightarrow \mathfrak{R}^{2n}$, so

$$a) \frac{e_L}{\sqrt{1 + C^T P_R C}} \in L_2 \cap L_\infty$$

$$b) \bar{\omega}_R \in L_\infty, \hat{\omega}_R \in L_2 \cap L_\infty$$

$$c) \quad \beta = \frac{\bar{\omega}_R \mathbf{C}}{1 + \|\mathbf{C}_T\|_\infty}, \quad \beta \in L_2 \cap L_\infty$$

where $\bar{\omega}_R = \hat{\omega}_R - \omega_R^*$. β can be considered a normalized error and \mathbf{e}_L is normalized by $\|\mathbf{C}_T\|_\infty$. β is included in L_2 . This assure that this gain converge to a small value when $t \rightarrow \infty$. By this way, the output error $\mathbf{e}_L \in L_2 \cap L_\infty$, $\mathbf{e}_L \rightarrow \mathbf{0}$ when $t \rightarrow \infty$. Moreover, the derivate of the parametric error $\dot{\bar{\omega}}_R \in L_2 \cap L_\infty$ and $\dot{\bar{\omega}}_R \rightarrow 0$ when $t \rightarrow \infty$.

Proof of Lemma 1: The proof of Lemma 1, that assure the convergence of $\bar{\omega}_R$, can be found in Bodson [2] theorem 2.4.4, and here will be omitted. □

When ω_{RM} is smaller than M_{R0} , $\sigma_R = 0$, the signal used in the control law feedback is ω_{RM} . In this case, the real control law parameters θ need to be known. To obtain them it is necessary to use a tachometer, as presented in section VII.

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