A MRLS Speed Estimator and RMRAC Applied to Encoderless Induction Motor Drives

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Abstract: - This paper proposes a Modified Recursive Least Squares (MRLS) with a Robust Model Reference Adaptive Controller (RMRAC). This structure is used to assure good performance in a wide speed range, including low and zero-speed conditions. Experimental results are given to show the effectiveness of the proposed controller.

Key-Words: - Speed Sensorless, Induction Motor Drive, Recursive Least Square, RMRAC.

1 Introduction

For a decade, induction motor 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 effort has been devoted the elimination of the speed sensor coupled to the shaft of the motor, presented in the conventional closed loop servo systems. The motivations for substitution of the speed sensor for estimation techniques is the cost, usually a speed sensor is a expensive component, and the 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 the stator voltages and currents. This technique is designed by two linear regression models derived from the machine electrical equations.

But all these techniques are based on back EMF measurement, and fail at low and zero speed because induced voltages are too low to be correctly. 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. So these methods are motor parameters independent and allow estimating the speed at low and zero stator frequency.

However these techniques can produce toque ripple and saturation in the main path and around the rotor slots causes an additional modulation which interferes in the rotor speed estimation.

To assure performance in a wide speed range, including low and zero speed conditions, this paper proposes a Modified Recursive Least Squares (MRLS) with a Robust Model Reference Adaptive Controller (RMRAC) for a speed sensorless induction motor drive. The MRLS is obtained modifying a Recursive Least Squares [12] using a sigma-modification. Moreover, a direct estimation technique of the rotor flux estimation is used to obtain an IFOC independent of the rotor time constant. The obtained controller is used to assure performance in a wide speed range, including low and zero speed conditions. Experimental results are presented to verify the dynamic performance of the resulting closed-loop system.

$$\begin{bmatrix} \dot{I}_{ds} \\ \dot{I}_{qs} \\ \dot{I}_{dr} \\ \dot{I}_{qr} \end{bmatrix} = \begin{bmatrix} -\frac{R_S}{\sigma.L_S} & \omega+\frac{\lambda_p.\omega_R.L_M^2}{\sigma.L_S.L_R} & \frac{R_R.L_M}{\sigma.L_S.L_R} & \frac{\lambda_p.\omega_R.L_M}{\sigma.L_S} \\ -\omega-\frac{\lambda_p.\omega_R.L_M^2}{\sigma.L_S.L_R} & -\frac{R_S}{\sigma.L_S} & -\frac{\lambda_p.\omega_R.L_M}{\sigma.L_S} & \frac{R_R.L_M}{\sigma.L_S.L_R} \\ \frac{R_S.L_M}{\sigma.L_S.L_R} & -\frac{\lambda_p.\omega_R L_M}{\sigma.L_R} & -\frac{R_R}{\sigma.L_R} & \omega-\frac{\lambda_p.\omega_R}{\sigma} \\ \frac{\lambda_p.\omega_R L_M}{\sigma.L_R} & \frac{R_S.L_M}{\sigma.L_S.L_R} & -\omega+\frac{\lambda_p.\omega_R}{\sigma} & -\frac{R_R}{\sigma.L_R} \end{bmatrix} \begin{bmatrix} I_{ds} \\ I_{qs} \\ I_{dr} \\ I_{qr} \end{bmatrix} + \begin{bmatrix} \frac{1}{\sigma.L_S} & 0 \\ 0 & \frac{1}{\sigma.L_S} \\ -\frac{L_M}{\sigma.L_S.L_R} & 0 \\ 0 & -\frac{L_M}{\sigma.L_S.L_R} \end{bmatrix} \begin{bmatrix} V_{ds} \\ V_{qs} \end{bmatrix} \qquad (1)$$

2 Problem Formulation

The *dq* model of two phase IM with the electrical variables referred to an arbitrary *dq* rotating frame is given by (1). The mechanical model is given by

$$T_E = \lambda_p.L_M\left(I_{dr}.I_{qs} - I_{ds}.I_{qr}\right) \qquad (2)$$

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

Where $\sigma = 1 - L_M{}^2/\left(L_S\,L_R\right)$, 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 stator fundamental and rotor slip frequencies. 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). λ_p is the number of pole pairs of the motor. The equation (1) was considered the motor perfectly balanced and regardless the saturation phenomena. The equation (2) represents the coupling between the electrical and mechanical models, which is given by the equation (3). By linearizing the electrical model of the motor (1)-(2) using the IFOC technique [11], it results in the following equations

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

$$T_E = \frac{\lambda_p\,L_M{}^2\,I_{qs}\,I_{ds}^*}{L_R} \qquad (5)$$

where I_{ds}^* is the reference current for the direct stator current (I_{ds}). This current is assumed constant to ensure a constant level of machine magnetization. This assumption is necessary to establish the IFOC.

The frequency of the voltage applied to the IM is

$$\omega = \lambda_p\,\omega_R + \left(R_R I_{qs}/L_R I_{ds}\right) \qquad (6)$$

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. Considering applications where the speed sensor can not be used to obtain ω_R, a estimation algorithm can be used and it is presented in next section.

2.1 Rotor Speed Estimation Algorithm

Consider the two phase induction motor defined in (1)-(3) and referred to the stator fixed frame,

$$\begin{bmatrix} V_{\alpha s} \\ V_{\beta s} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} a_1 & 0 & p.L_M & 0 \\ 0 & a_1 & 0 & p.L_M \\ p.L_M & \omega_R L_M & a_2 & \omega_R L_M \\ -\omega_R L_M & p L_M & -\omega_R L_M & a_2 \end{bmatrix} \begin{bmatrix} I_{\alpha s} \\ I_{\beta s} \\ I_{\alpha r} \\ I_{\beta r} \end{bmatrix} \qquad (7)$$

where $a_1 = R_S + p\,L_S$, $a_2 = R_R + p\,L_R$ and p (= d/dt). Using the equation (7) it is possible to reformulate the problem of estimating the speed as a problem of estimating the parameters based on a linear regression model. Linear regression models have the structure $\mathbf{Y} = \mathbf{C}\,\hat{\omega}_R$ where, considering the rotor speed ω_R as the only unknown parameter, **Y** and **C** are given by (8).

$$\mathrm{Y} = \begin{bmatrix} \ddot{I}_{\alpha s} + P_1\dot{I}_{\alpha s} + \frac{R_S R_R}{L_S L_R\,\sigma} I_{\alpha s} - \frac{1}{\sigma L_S}\dot{V}_{\alpha s} - \frac{R_R}{L_S L_R \sigma} V_{\alpha s} \\ \ddot{I}_{\beta s} + P_1\dot{I}_{\beta s} + \frac{R_S R_R}{L_S L_R \sigma} I_{\beta s} - \frac{1}{\sigma L_S}\dot{V}_{\beta s} - \frac{R_R}{L_S L_R \sigma} V_{\beta s} \end{bmatrix}$$

$$\mathrm{C} = \begin{bmatrix} -\dot{I}_{\beta s} - \frac{R_S}{L_S\,\sigma} I_{\beta s} + \frac{1}{L_S\,\sigma} V_{\beta s} \\ \dot{I}_{\alpha s} + \frac{R_S}{L_S\,\sigma} I_{\alpha s} - \frac{1}{L_S\,\sigma} V_{\alpha s} \end{bmatrix} \qquad (8)$$

Let us assume that the derivatives presented in (8) are measurable quantities. In the implementation these quantities are obtained by state variable filters (SVF) [12]. These filters are developed by discretization of transfer function given by,

$$\frac{V_{f\alpha s}}{V_{\alpha s}} = \frac{V_{f\beta s}}{V_{\beta s}} = \frac{I_{f\alpha s}}{I_{\alpha s}} = \frac{I_{f\beta s}}{I_{\beta s}} = G_f(s) = \frac{\omega_C{}^3}{\left(s+\omega_C{}^3\right)^3} \qquad (9)$$

where $\omega_C = 5\,\omega$ and inputs $V_{\alpha s}$, $V_{\beta s}$, $I_{\alpha s}$ and $I_{\beta s}$ are used to obtain the filtered signals $V_{f\alpha s}$, $V_{f\beta s}$, $I_{f\alpha s}$ and $I_{f\beta s}$.

In very low speed the conventional RLS used in [12] presents a poor performance, because of the low value of the FCEM in this case.

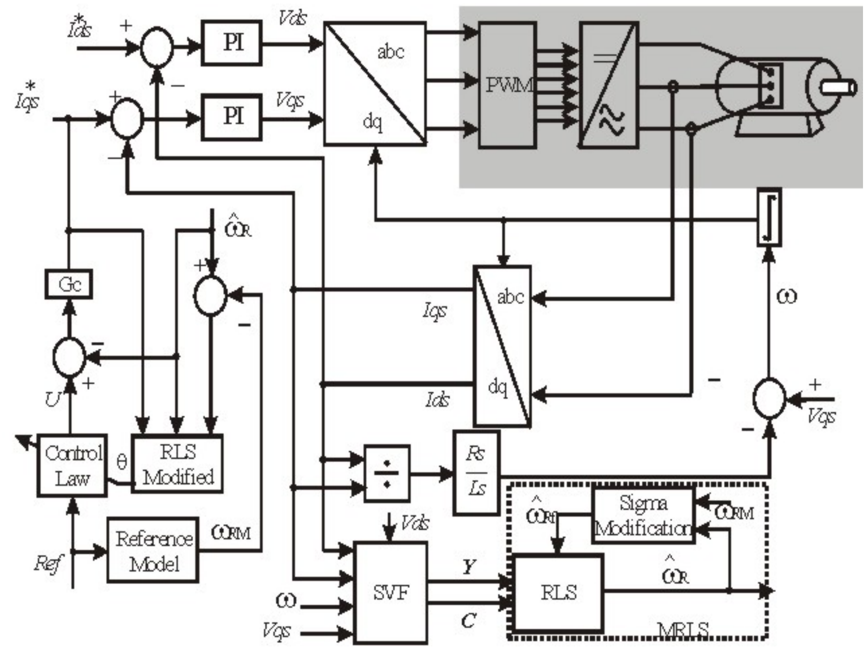

Fig. 1– Controller Structure

3 Problem Solution

To overcome this difficulty, a sigma modification was applied to the RLS algorithm to obtain a Modified Recursive Least-Square (MRLS) given by

$$\hat{\omega}_{\mathrm{Rf}} = \sigma_{\mathrm{R}}\, \hat{\omega}_{\mathrm{R}} + \left(1-\sigma_{\mathrm{R}}\right)\omega_{\mathrm{RM}} \tag{10}$$

where

$$\sigma_R = \begin{cases} 0 & if \ \ \|\omega_{\mathrm{RM}}\| < M_{R0} \\ \sigma_{R0}\left(\dfrac{\|\omega_{\mathrm{RM}}\|}{M_{R0}} - 1\right) & if \ \ M_{R0} \le \|\omega_{\mathrm{RM}}\| < 2M_{R0} \\ \sigma_{R0} & if \ \ \|\omega_{\mathrm{RM}}\| \ge 2M_{R0} \end{cases} \tag{11}$$

$$\hat{\omega}_R(t) = \hat{\omega}_{Rf}(t-1) + \boldsymbol{K}(t-1)\boldsymbol{e}(t) \tag{12}$$

$$\boldsymbol{e}(t) = \boldsymbol{Y}(t) - \boldsymbol{C}(t-1)\hat{\omega}_{Rf}(t-1) \tag{13}$$

$$\boldsymbol{K}(t) = \frac{P_R(t-1)\boldsymbol{C}^T(t)}{1+\boldsymbol{C}^T(t)P_R(t-1)\boldsymbol{C}(t)} \tag{14}$$

$$P_R(t) = P_R(t-1) + \boldsymbol{K}(t)\boldsymbol{C}(t)P_R(t-1) \tag{15}$$

ω_{RM} is a model reference output obtained from the control law structure. It will be presented in the next section.

The algorithm convergence is proved by the Lemma 1, which it is presented in the appendix. By this algorithm, it is possible to obtain the estimate speed $\hat{\omega}_R$, that it will be used on the control law.

3.1 Controller Structure

The proposed sensorless speed control structure is shown in the Fig.1. The currents signals I_{ds} and I_{qs} are obtained from the machine supply currents and compared with the I^*_{ds} and I^*_{qs} reference currents. A RMRAC control law is used to generate the I^*_{qs} current by the difference between the estimated speed $\hat{\omega}_R$ and the output of the reference model ω_{RM} [9]. The adaptive RMRAC control law was used to simplify and to reduce the time design. Besides this, it makes possible to project the mechanical controller without the exact knowledge of the plant (Induction Motor/Inverter system). Moreover, the speed estimator algorithm with RMRAC controller results in a sensorless scheme tolerant to unmodeled dynamics that they can be present in the overall system.

The RMRAC structure is obtained considering a SISO plant (Single-Input Single-Output),

$$\hat{\omega}_R = G(s)U \tag{16}$$

$$G(s) = G_0(s)\left[1+\mu\,\Delta_{\mathrm{m}}(s)\right] + \mu\,\Delta_{\mathrm{a}}(s) \tag{17}$$

where $G(s)$ is the plant transfer function, $G_0(s)$ is the modeled part of the plant and $\mu\Delta_{\mathrm{m}}(s)$ and $\mu\Delta_{\mathrm{a}}(s)$ are additive and multiplicative perturbations, respectively. The plant $G_0(s)$ is a strictly proper transfer function, such as $G_0(s) = k_{\mathrm{p}}\, Z_0(s)/R_0(s)$, where $Z_0(s)$ and $R_0(s)$ are monic polynomials with m and n degree, 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 a monic Hurvitz polynomial of degree $m \le (n-1)$;

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;

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

$$\omega_{RM} = G_M(s)\,Ref = \left(k_m Z_m(s)/R_m(s)\right)Ref \qquad (18)$$

where $G_M(s)$ has a relative degree $n^* = n - m$, $Z_m(s)$ and $R_m(s)$ are Hurwitz polynomials, *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, in despite of the disturbances $\Delta_a(s)$ e $\Delta_m(s)$, satisfying S3.

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

$$\dot{w}_1 = F\,w_1 + qU \qquad (19)$$

$$\dot{w}_2 = F\,w_2 + q\hat{\omega}_R \qquad (20)$$

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

$$U = -\left(w_1\,\theta_1^T + w_2\,\theta_2^T + \hat{\omega}_R\theta_3 + Ref\right)\Big/\theta_4 \qquad (21)$$

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

3.2 RMRAC Parameters Estimation Algorithm

The control law parameters are obtained by

$$\dot{\boldsymbol{\theta}} = -\sigma\boldsymbol{P\theta} - \frac{\varepsilon_1\boldsymbol{P\zeta}}{\bar{m}} \qquad (22)$$

$$\dot{\boldsymbol{P}} = -\frac{\boldsymbol{P\zeta\zeta}^{\mathrm{T}}\boldsymbol{P}}{\bar{m}} + \left(\lambda\boldsymbol{P} - \frac{\boldsymbol{P}^2}{R^2}\right)\bar{\mu}^{2} \qquad (23)$$

and $\boldsymbol{P} = \boldsymbol{P}^{\mathrm{T}}$ is so that

$$0 < \boldsymbol{P}(0) \le \lambda R^2\boldsymbol{I},\;\; \mu^2 \le k_\mu\,\bar{\mu}^2 \qquad (24)$$

$$\bar{m} = 1 + \alpha_1\left[m\right]^2,\quad \boldsymbol{\zeta} = G_m\,\boldsymbol{I}\,\boldsymbol{w} \qquad (25)$$

$$\dot{m} = \delta_0 m + \delta_1\left(|U| + |\hat{\omega}_R| + 1\right)$$
$$m(0) > \frac{\delta_1}{\delta_0},\; \delta_1 \ge 1 \qquad (26)$$

where α_1, δ_0, δ_1, δ_2, λ, $\bar{\mu}$ and R^2 are positive constants and δ_0 satisfies $\delta_0 + \delta_2 \le \min[p_0, q_0]$. $q_0 \in \Re_+$ is such that the $G_M(s-q_0)$ poles and the $(F + q_0\mathrm{I})$ eigenvalues are stable. The sigma modification σ in (22) is given by

$$\sigma = \begin{cases} 0 & if\;\; \|\theta\| < M_0 \\ \sigma_0\left(\dfrac{\|\theta\|}{M_0} - 1\right) & if\;\; M_0 \le \|\theta\| < 2M_0 \\ \sigma_0 & if\;\; \|\theta\| \ge 2M_0 \end{cases} \qquad (27)$$

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

$$\varepsilon_1 = e_1 + \boldsymbol{\theta}^{\mathrm{T}}\boldsymbol{\zeta} - G_m\boldsymbol{\theta}^{\mathrm{T}}\mathbf{w} \qquad (28)$$

The convergence of this algorithm is described by the theorem 1, presented in the appendix.

4 Experimental Results

The RMRAC sensorless speed servo was implemented in a PC-based platform driving an induction motor. The motor is a Y-connected two-pole, 0.9 Hp, 3500 rot/min, 380-V/2.7-A type. Motor parameters were obtained by no-load test, locked-rotor test and board data, and are presented in TABLE1.

The design is begun by the definition of the reference model, once it imposes the required closed-loop dynamic to the plant. The following reference model was used.

$$G_M(s) = \frac{8.2\,s + 0.01}{s^2 + 8.2\,s + 0.01} \qquad (1)$$

The existence of a large difference between the dynamic of the model reference, and the dynamic of the plant, may cause problems in the RMRAC controller. To overcome this problem a pre-compensator $G_C(s)$ is used.

$$G_C(s) = \frac{0.009\,s + 0.01}{s} \qquad (2)$$

The gains F and q used in (19) and (20) are 10 and 1, respectively. The DC bus was limited in 177V/5A, and the sampling time used is 555 µs. The reference signal was kept at zero until 4s because of the machine magnetization.

Figs. 4, 5 and 6 show the speed sensorless controller responses obtained using the reference presented in Fig. 3. In a first test, the system was initially operating at 0 rad/s, when the speed reference was increased until 60 rad/s. After 10 sec, the speed reference is reduced until -30 rad/s. It can be seen that some spikes occur in the speed estimated, during low speed. These spikes cause some oscillations at d-axis and q-axis currents, as shown in Fig. 6. Note that these oscillations do not degrade the control action.

Figs. 8, 9 and 10 show the proposed controller responses, using the reference presented in Fig. 7. The system was initially operating at 0 rad/s, when the speed reference was 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. Fig. 5 presents the control law parameters obtained and Fig.6 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 was obtained. These results are obtained with no load.

For the load step test, a DC generator was connected to the IM rotor, the parameters of which are given in TABLE 2. The system was initially operating at 0 rad/s, when the speed reference was increased to 60 rad/s. During the test, a 15 Ω resistance was connected on the generator terminals in 20 sec. Fig. 11 shows the performance in an instantaneous load torque change, and Fig. 12 presents the quadrature current obtained.

TABLE 1
Motor Parameters

Power	0.9 Hp
Nominal Speed	3400 RPM
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

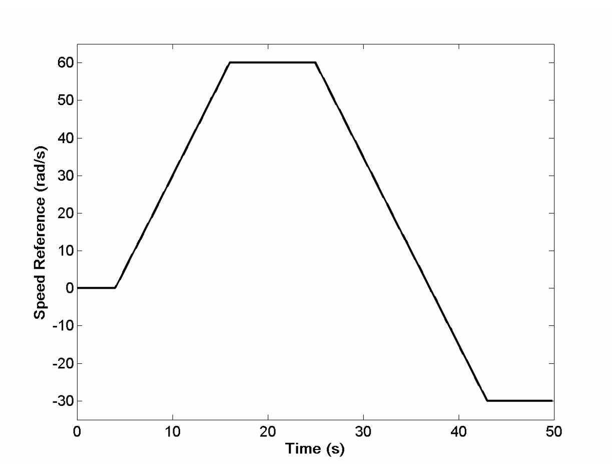

Fig. 2– Reference signal of the first test

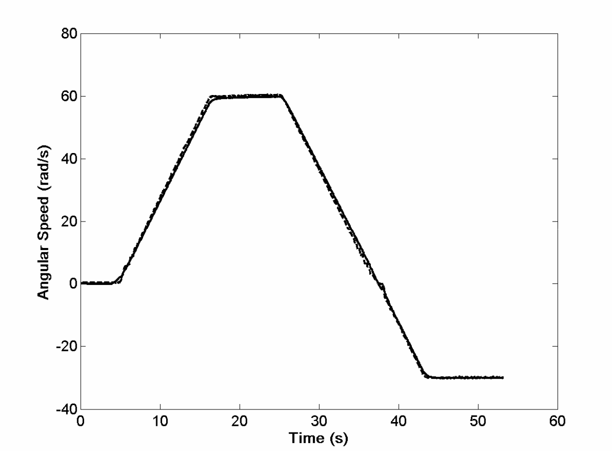

Fig. 3– Estimated speed (solid line)and measured speed (dashed Line) in first test

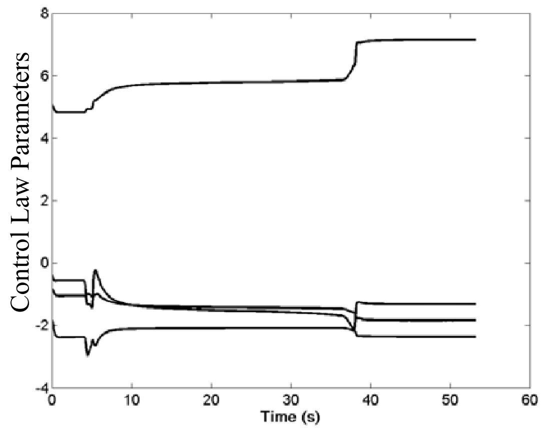

Fig. 4– Control law parameters during the first test

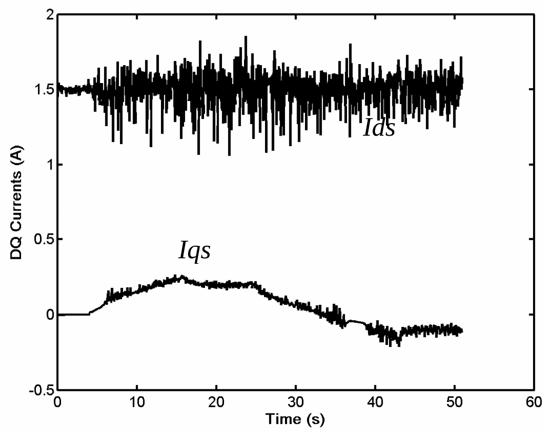

Fig. 5– Quadrature current obtained during the first test

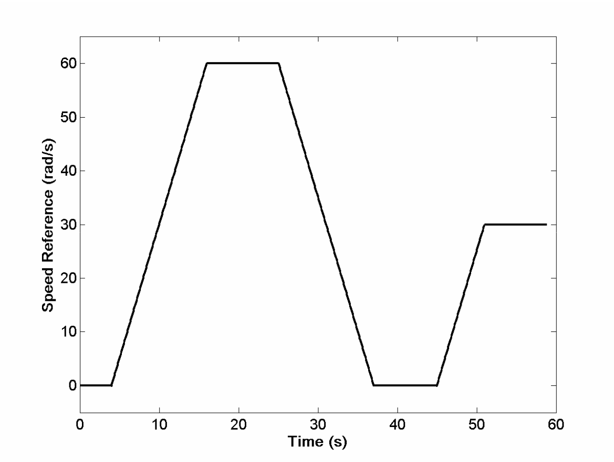

Fig. 6– Reference signal in the second test

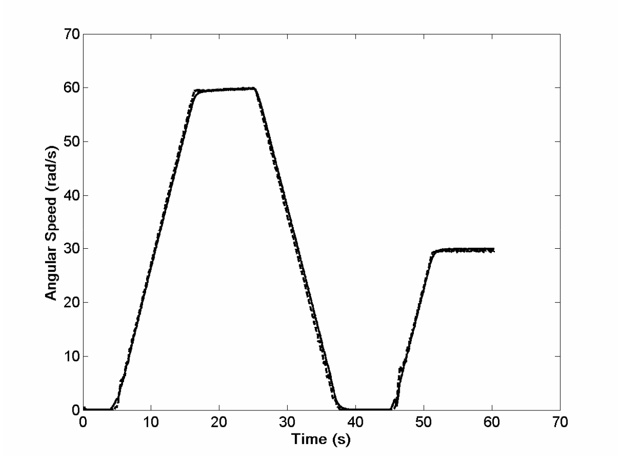

Fig. 7– Estimated speed (solid line)and measured speed (Dashed Line) in second test

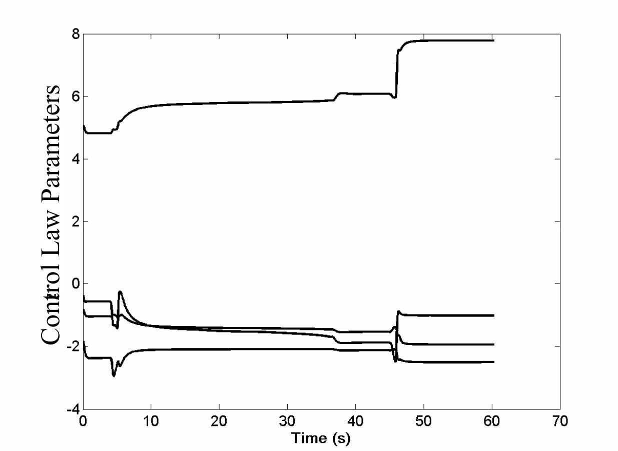

Fig. 8– Control law parameters during the second test

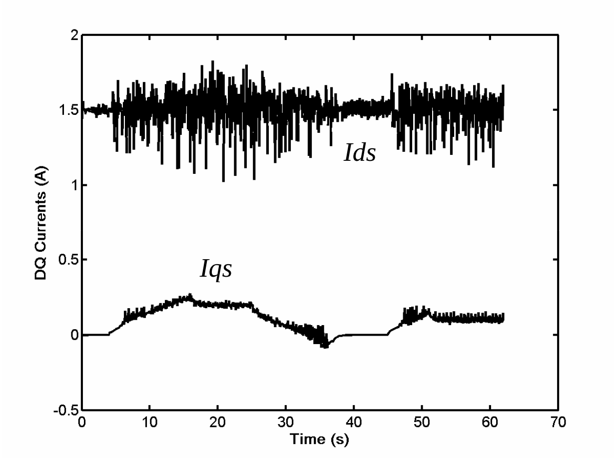

Fig. 9– Quadrature current obtained during the second test

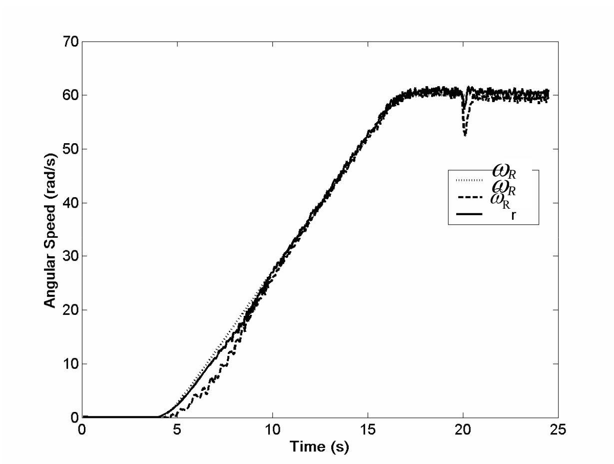

Fig. 10– Estimated speed (Dashed Line), measured speed (solid line) and model reference speed (dotted line) in torque disturbance test

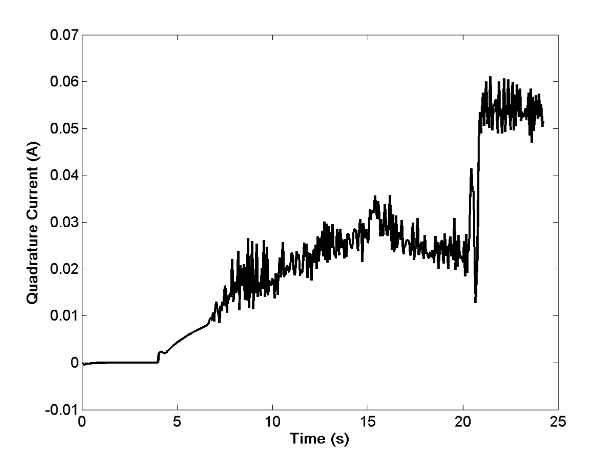

Fig. 11– Quadrature current obtained during the torque disturbance test

5 Conclusion

This paper describes a Modified Recursive Least Squares (MRLS) with a Robust Model Reference Adaptive Controller (RMRAC) applied to a speed sensorless servo system using three-phase induction motor. A speed estimator is incorporated to the control system to avoid the use of mechanical sensors. The experimental results demonstrate the effectiveness of the proposed control scheme, including the control at low speed and compensation of torque disturbances.

Moreover, this 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 measures of the plant output signal to control it, but uses an observed signal $\hat{\omega}_R$.

Appendix:

Lemma 1: Let $\boldsymbol{C}(t)$ given by (8) where $V_{\alpha s}$, $V_{\beta s}$, $I_{\alpha s}$, $I_{\beta s}$ and them derivates are piecewise continuous function of time Moreover, consider the linear error equation $\boldsymbol{e}_L = \left(\hat{\omega}_R - \overset{*}{\omega}_R\right)\boldsymbol{C}(t)$ (vide [2] eq. 2.4.3) with a RLS algorithm presented in (10)-(15) and $\sigma_R = 1$. Defining a vector $\boldsymbol{C}$: $\Re_+ \rightarrow \Re^{2n}$, so

a) $\dfrac{\boldsymbol{e}_L}{\sqrt{1+\boldsymbol{C}^T \boldsymbol{P}_R \boldsymbol{C}}} \in L_2 \cap L_\infty$

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

c) $\beta = \dfrac{\bar{\omega}_R \boldsymbol{C}}{1+\left\|\boldsymbol{C}_T\right\|_\infty}$, $\beta \in L_2 \cap L_\infty$

where $\bar{\omega}_R = \hat{\omega}_R - \overset{*}{\omega}_R$. β can be considered a normalized error and $\boldsymbol{e}_L$ is normalized by $\left\|\boldsymbol{C}_T\right\|_\infty$. 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 $\boldsymbol{e}_L \in L_2 \cap L_\infty, \boldsymbol{e}_L \rightarrow 0$ 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.

□

Theorem 1: Assume that $\hat{\omega}_R$ satisfy the Lemma 1 and Ref and $\dot{Ref}$ bounded. So, there is a vector $\boldsymbol{\theta}$ so that all the signals in the feedback system by the process (16), with the controller (18)-(21) and the parametric control law adaptation (22)-(28), together with the speed estimation algorithm (10)-(15), are limited for all initial condition. By this way, there is a constant $\gamma_1 > 0$ and a $\bar{\varepsilon}$ so that the tracking error $e_1 = \hat{\omega}_R - \omega_{RM}$ belongs to a residual set

$$D_e = \left\{ e_1 : \lim_{T\rightarrow\infty} \sup_{T\rightarrow\infty} \frac{1}{T}\int_{t_0}^{t_0-T} e_1(\tau)\,d\tau, \forall t_0 \geq 0, T > 0 \right\}$$

Moreover, in the absence of modeling error, the adaptive control law algorithm guarantees boundedness of all the signals as well as convergence of the tracking error e_1 to zero.

Proof of Theorem 1: The proof of the Theorem 1, that it guarantees the convergence of e_1, can be found in Leal [10], and here will be omitted.

□

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