A DECENTRALIZED RMRAC SCHEME FOR CARRIER-BASED STATOR CURRENT CONTROL OF INDUCTION MOTORS

H. T. Câmara, L. Michels, J. R. Pinheiro, H. L. Hey,H. Pinheiro and H. A. Gründling
Power Electronics and Control Research Group – GEPOC
Federal University of Santa Maria – UFSM
CEP: 97105-900 – Santa Maria, RS – Brazil
htcamara@hotmail.com, michels@ieee.org - http://www.ufsm.br/gepoc/

Abstract **–This paper presents a decentralized robust model reference adaptive control (RMRAC) scheme applied to the dynamic decoupling of the stator currents of a carrier-based modulated induction motor. The objective of the proposed controller structure is to overcome the problems of dynamic performance and robustness of others adaptive techniques for this application. The proposed technique controls independently the d-axis and q-axis using a PI pre-compensator allied to a direct RMRAC SISO controller in each axis. The adaptive scheme provides a direct tuning of the controller parameters while the controller structure ensures the robustness. Toward demonstrate the robustness and dynamic performance of the proposed controller under different operations conditions are presented experimental results for a speed servomechanism driven by a 3/4HP induction motor.**

KEYWORDS

AC Drives, Robust Control, Dynamic Decoupling.

I. INTRODUCTION

Industrial applications such as robotics, rolling mills and machine tools demands fast and precise response of torque, speed and position servo drives. Since a long time ago the DC machines have been used for these applications due to their simple control dynamics. However, these machines are inferior to the squirrel-cage induction machines in most others respects, such as cost, power, weight, size, ruggedness, maintenance requirements and maximum speed. These advantages, allied to the development of the power electronics, microelectronics and vector control theory, disseminated the induction machines as the foundation for electrical servo drive systems in last decades.

Toward make possible the control of induction machines were developed several vector control schemes in order to linearize the complex dynamic behavior of them, concurrently with the development of power static converters technology. Among the vector control schemes and drives topologies developed, the indirect field-oriented control (IFOC) driven by voltage-source inverters (VSI) became an industry standard for high dynamic performance induction motors drives in low and medium power applications, due to it simplicity, high reliability and fast response.

Besides of their advantages, to obtain a high performance in IFOC-VSI drives is necessary: i) to preserve a correct tuning of the field orientation, guaranteed by the correct determination of the slip frequency; ii) a accurate control of the stator currents. The slip frequency may be correctly determined using an algorithm for the calculation of the electrical time-constant of the rotor circuit, which can be done using different techniques [1]. For the control of the stator currents in VSI-induction motors drives many techniques have been developed. Among them, it can be highlighted the hysteresis-based controllers and the carrier-based controllers [2]. The hysteresis-based current control forces the currents to follows a reference using variable frequency of operation. The carrier-based one determines the voltages that must feed the motor to produce the reference currents. Due to the carrier-based current control operates with fixed switching frequency, it presents low ripple currents and have well-defined harmonic spectrum characteristics, as results it is widely utilized.

The carrier-based current control adds complexity to the model of the motor due to the inclusion of the dynamics of the stator circuit. The resulting model presents a significant cross coupling between the d and q axis, where the coupling level is variable and directly proportional to the frequency of the voltage that feed the motor, been more prominent when the motor is operated above the nominal speed [3]. In order to guarantee a fast response of the carrier-based controller in a wide speed range is required a suitable controller with a decoupling compensation scheme. In addition to the cross coupling dynamics presents in the carrier-based current control, the electrical model of the induction motor using a carrier-based current control presents a higher sensibility to parameters variations of the motor parameters. These parameters may change considerably during the operation due to the changes in the temperature of the motor and due to saturation phenomena [1]. The coupling level and the variation of the stator circuit parameters may causes detuning of the carrier-based current controllers, resulting in a poor dynamic response.

Intending to overcome the poor response due to detuning of the carrier-based current control schemes were proposed the use of several different control structures and the employment of on-line tuning methods to ensure a good dynamic response even when the plant suffer considerable parametric variations. In order to maintain well tuned the different controllers were proposed on-line tuning algorithms based basically on estimation of parameters [3] and adaptive controllers [4]. The main drawbacks of these methods are: i) the requirement of signals persistently exciting (PE) in the reference input and a well known dynamics of the plant to ensure convergence to the real parameters; ii) the adaptive controllers can present the undesirable bursting phenomena

[5],[6] if the input reference signals are not PE, resulting in an unacceptable transitory behavior; iii) the existence of unmodeled dynamics can become adaptive controllers unstable even for small external disturbs [6]. It is well known that the model of the motor presents other dynamics that usually are unconsidered. As almost all the adaptive controllers for IFOC are based on the linearized model of the machine, there are not guarantee of robustness for these controllers [2]-[4]. Aiming to overcome the robustness problem was proposes the robust model reference adaptive controller [8], which presents robustness even if the plant presents unstructured dynamics [7]. These controller also presents the parameter adaptation algorithm that does not require signals PE in the reference input and presents absolute convergence to the real values in the absence of unstructured dynamics and a bounded error when exist unstructured dynamics. [7]

This paper proposes the use of a decentralized RMRAC controller for the dynamic decoupling of carrier-based induction motor IFOC schemes, in order to overcome the robustness problems of the others adaptive controllers for IFOC previously reported. The decentralized RMRAC controller had already been used for dynamic decoupling of a weakly coupled plant with good results [9]. Although exist a genuine MIMO RMRAC controller [10] presenting better dynamic behavior, the decentralized RMRAC was chase due to the smaller computational demanding.

The proposed technique controls independently the d-axis and q-axis using a PI pre-compensator allied to a direct RMRAC SISO controller [8] in each axis. The adaptive scheme provides a direct tuning of the controller parameters while the controller structure ensures the robustness. The tolerance of the control law to unmodeled dynamics permit to consider the cross coupling as an unmodeled dynamic for the controller. In this way, since the plant presents a weak-coupling effect, the controllers must to support and compensate the coupling dynamics.

This paper is organized as follows: the Section II contains the description of the plant while in the Section III presents the proposed carrier-based IFOC stator current control structure. The proposed RMRAC control law and the parameter adaptation algorithm are described in the Section IV. The Section V presents experimental results using a 3/4HP induction motor used to verify performance of the proposed controller.

II. PLANT DESCRIPTION

The model of plant composed by the three-phase induction motor employing the carrier-based current control and the mechanical load can be described by (1), (2), and (3),

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

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

where $\sigma = 1 - L_M{}^2 / (L_S L_R)$, V_{ds}, V_{qs} are the stator voltages in an arbitrary dq rotating frame. 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 in an arbitrary dq rotating frame. 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). λ are the number of pole pairs of the motor.

The equation (1) gives a state-space representation of the electrical model of the induction motors in an arbitrary rotating reference frame, considering 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 given by (1) and (2) using the IFOC technique results in the following equations

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

$$T_E = \frac{\lambda L_M{}^2 \, I_{qs} \, I_{ds}{}^*}{L_R} \tag{5}$$

where $I_{ds}{}^*$ is the reference current imposed to I_{ds}, assumed to be constant to ensure a constant level of machine magnetization. This assumption is necessary to obtain a linearization in (5). However, for the linearization of the induction motor model to be valid, is necessary to impose correctly the frequency of the voltage in the motor. This frequency can be determined by using

$$\omega = \lambda \, \omega_r + \frac{R_R I_{qs}}{L_R I_{ds}} \tag{6}$$

$$\begin{bmatrix} \dot{I}_{ds} \\ \dot{I}_{qs} \\ \dot{I}_{dr} \\ \dot{I}_{qr} \end{bmatrix} = \begin{bmatrix} -\dfrac{R_S}{\sigma . L_S} & \omega + \dfrac{\lambda . E_R . L_M{}^2}{\sigma . L_S . L_R} & \dfrac{R_R . L_M}{\sigma . L_S . L_R} & \dfrac{\lambda . \omega_R . L_M}{\sigma . L_S} \\ -\omega - \dfrac{\lambda . \omega_R . L_M{}^2}{\sigma . L_S . L_R} & -\dfrac{R_S}{\sigma . L_S} & -\dfrac{\lambda . \omega_R . L_M}{\sigma . L_S} & \dfrac{R_R . L_M}{\sigma . L_S . L_R} \\ \dfrac{R_S . L_M}{\sigma . L_S . L_R} & -\dfrac{\lambda . \omega_R L_M}{\sigma . L_R} & -\dfrac{R_R}{\sigma . L_R} & \omega - \dfrac{\lambda . \omega_R}{\sigma} \\ \dfrac{\lambda . \omega_R L_M}{\sigma . L_R} & \dfrac{R_S . L_M}{\sigma . L_S . L_R} & -\omega + \dfrac{\lambda . \omega_R}{\sigma} & -\dfrac{R_R}{\sigma . L_R} \end{bmatrix} \begin{bmatrix} I_{ds} \\ I_{qs} \\ I_{dr} \\ I_{qr} \end{bmatrix} + \begin{bmatrix} \dfrac{1}{\sigma . L_S} & 0 \\ 0 & \dfrac{1}{\sigma . L_S} \\ -\dfrac{L_M}{\sigma . L_S . L_R} & 0 \\ 0 & -\dfrac{L_M}{\sigma . L_S . L_R} \end{bmatrix} \begin{bmatrix} V_{ds} \\ V_{qs} \end{bmatrix} \tag{1}$$

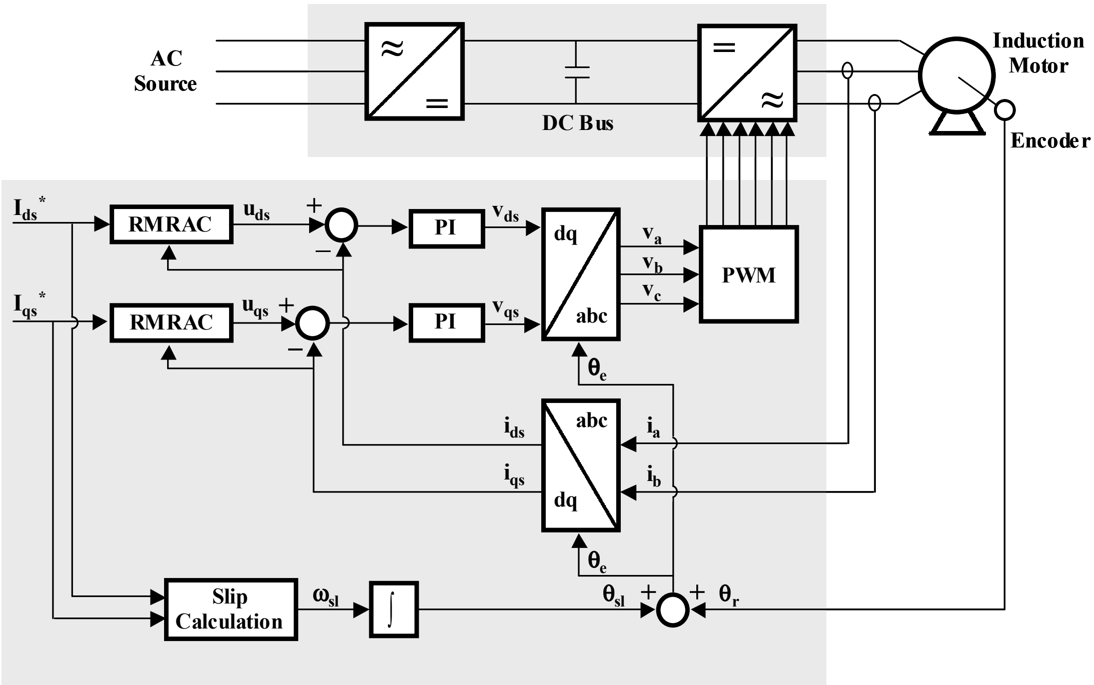

Fig. 1. Proposed Carrier-Based IFOC Control Structure

III. PROPOSED CARRIER-BASED IFOC STATOR CURRENT CONTROL STRUCTURE

The proposed stator current control structure is shown in the Fig.1. The control considers the *d* and *q*-axis as two independent SISO plants, does not requiring to taking into account the existence of cross coupling.

In order to obtain a null error in the *dq*-axis current references in steady-state operation are used proportional-integral (PI) pre-compensators in the *d* and *q*-axis. In addition, two RMRAC SISO controllers are used to compensate the detuning of the PI controllers due to cross coupling effect and parametric variations. Besides the controller does not have a proper structure to compensate the coupling effects, the RMRAC controllers can compensate these dynamics if them are not much prominent. This is possible due to the tolerance of the control law to unmodeled dynamics that permit to consider the cross coupling as an unmodeled dynamic for the controller.

The RMRAC control law u is obtained as follow in the Section IV.

IV. RMRAC CONTROLLER DESCRIPTION

1) RMRAC Controller Structure: Considering a SISO plant (single-input single-output), given as

$$\begin{aligned} y &= G(s)\,u \\ &= \{G_0(s)[1+\mu\Delta_m(s)]+\mu\Delta_m(s)\}u \end{aligned} \tag{7}$$

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

$$G_0(s) = k_P \frac{Z_0(s)}{R_0(s)} \tag{8}$$

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. $Z_0(s)$ is a monic Hurvitz polynomial of degree $m \le (n-1)$.

S2. $R_0(s)$ is a monic polynomial of degree n.

S3. The signal of k_P and the values of m and n are known.

Furthermore, for the unmodeled plant part is assumed that:

S4. $\Delta_a(s)$ is a strictly proper stable transfer function;

S5. $\Delta_m(s)$ is a stable transfer function;

S6. A lower 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

$$y_m = W_m(\mathrm{s})r = K_m \frac{N_m(\mathrm{s})}{D_m(\mathrm{s})} r \tag{9}$$

where $D_m(s)$ is a Hurwitz polynomial of degree $n^* = n - m$ and $r(t)$ 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 y tracks the reference model output y_m as closely as possible, in despite of the disturbances $\Delta_a(s)$ e $\Delta_m(s)$, satisfying **S4 – S6.**

The plant input u and output y are used to generate n-1 dimensional auxiliary vectors, as in [9], so that

$$\dot{w}_1 = \mathbf{F}\,w_1 + q\,u \tag{10}$$

$$\dot{w}_2 = \mathbf{F}\,w_2 + q\,y \tag{11}$$

where $\mathbf{F}$ is a stable matrix and the $(\mathbf{F},q)$ is a controllable pair.

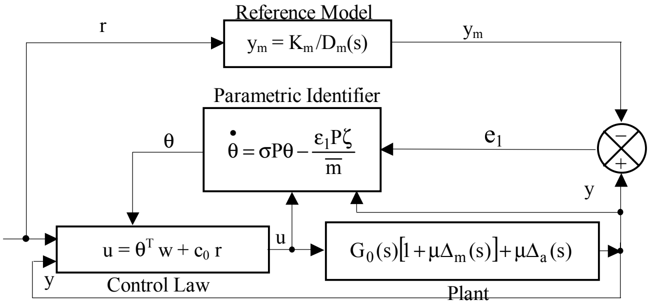

Fig. 2. Robust model reference adaptive control structure

The plant input u is taken as,

$$u = \boldsymbol{\theta}^{\mathrm{T}}\mathbf{w} + c_0\, r \tag{12}$$

where $\boldsymbol{\theta}^{\mathrm{T}} = [\theta_1^{\mathrm{T}}, \theta_2^{\mathrm{T}}, \theta_3]$ is a $(2n\text{-}1)$ dimensional vector of control parameters, c_0 is a scalar gain, and $\boldsymbol{w}^T = \lfloor w_1{}^T, w_2{}^T, y \rfloor$. The tracking error is $e_1 = y - y_m$.

2) Parameter Adaptation Algorithm:

A significant number of parameter estimation techniques have been successfully applied to identification problems [9]. In the proposed controller, a modified recursive least-squares (RLS) algorithm is used, which presents fast convergence, when compared with gradient algorithms. This basic RMRAC algorithm structure is illustrated in the Fig. 2.

The parameter adaptation algorithm is described as follows

$$\dot{\boldsymbol{\theta}} = \sigma \mathbf{P}\boldsymbol{\theta} - \frac{\varepsilon_1 \mathbf{P}\boldsymbol{\zeta}}{\bar{m}} \tag{13}$$

$$\dot{\mathbf{P}} = \lambda \bar{\mu}^{-2}\mathbf{P} - \left(\frac{\mathbf{P}\boldsymbol{\zeta}\boldsymbol{\zeta}^{\mathrm{T}}\mathbf{P}}{\bar{m}} + \bar{\mu}^2 \frac{\mathbf{P}^2}{R^2} \right) \tag{14}$$

where, $\mathbf{P} = \mathbf{P}^{\mathrm{T}}$ is so that

$$\begin{gathered} 0 < \mathbf{P}(0) \le \lambda R^2 I, \ \ \mu^2 \le k_\mu \bar{\mu}^2 \\ \bar{m} = 1 + \alpha_1 [m]^2, \quad \boldsymbol{\zeta} = \mathbf{W}_m \mathbf{I}\, \omega \end{gathered} \tag{15}$$

$$\dot{m} = \delta_0 m + \ \delta_1 \left(|u| + |y| + 1 \right),\ m(0) > \frac{\delta_1}{\delta_0},\ \delta_1 \ge 1 \tag{16}$$

where α, δ_0, δ_1, 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 poles of $W_m(s\text{-}q_0)$ and the eigenvalues of $(\mathbf{F} + q_0\mathbf{I})$ are stable and $\delta_2 \ge 0$. $p_0 > 0$ is defined in **S6** and σ in (17) can be expressed as

$$\sigma = \begin{cases} 0 & if \ \ \|\boldsymbol{\theta}\| < M_0 \\ \sigma_0 \left(\dfrac{\|\boldsymbol{\theta}\|}{M_0} - 1 \right) & if \ \ M_0 \le \|\boldsymbol{\theta}\| < 2M_0 \\ \sigma_0 & if \ \ \|\boldsymbol{\theta}\| \ge 2M_0 \end{cases} \tag{17}$$

where $M_0 > \|\boldsymbol{\theta}^*\|$ and $\sigma_0 > 2\bar{\mu}^2 / R^2 \in \Re^+$ are design parameters. As defined in [5], the modified error is given by

$$\varepsilon_1 = e_1 + \boldsymbol{\theta}^{\mathrm{T}}\boldsymbol{\zeta} - W_m \boldsymbol{\theta}^{\mathrm{T}}\mathbf{w} \tag{18}$$

TABLE I
Parameters of the plant

Mecânicos	Elétricos
J = 0.089 Kg.m^2	L_s = 0.2979 H
B = 0.00114 Kg.m^2/A	L_m = 0. 2763 H
Nominal Speed = 865RPM	L_r = 0. 3087 H
Poles = 8	R_s = 8.95 Ω
Power = 3/4 c.v	R_r = 7.50 Ω

or

$$\varepsilon_1 = \boldsymbol{\theta}^{\mathrm{T}}\boldsymbol{\zeta} + \mu\eta \tag{19}$$

V. EXPERIMENTAL RESULTS

The current control scheme presented in the Section III and IV was implemented in a PC-based platform driving an induction motor. This section presents experimental results using a PC-based platform using the motor whose parameters are given in Table I. These parameters were obtained by No-load test, Locked-rotor test and board data.

The design started defining the reference models, once they impose the closed-loop dynamic of the plant in model reference control laws. The following reference models were used the following reference models for the d and q-axis, respectively:

$$Wm_d(s) = \frac{100s + 10}{s^2 + 100s + 10} \tag{20}$$

$$Wm_q(s) = \frac{50s + 5}{s^2 + 50s + 5} \tag{21}$$

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. Toward overcome this problem are used two pre-compensators, been one in the d-axis and other in the q-axis.

$$PI_d(s) = \frac{5.06s + 505.99}{s} \tag{22}$$

$$PI_q(s) = \frac{3.795s + 284.622}{s} \tag{23}$$

The pre-compensator adopted was the PI once it presents error null in steady-state operation for constant disturbs. The gains of the PI compensator are calculated such that the compensated plant has a similar dynamic as the model reference required.

Additional an external PI a speed servo controller was applied to the mechanical part of the plant, as presented in fig.3. It is aiming to verify the capability of the proposed controller to compensate the cross coupling, more prominent at high speeds. By the error signal between speed reference W_{ref} and angular speed W_r the current I_{qs}^* is obtained. The gains were obtained using the algorithm in [12], and with parameters shown in Table 1.

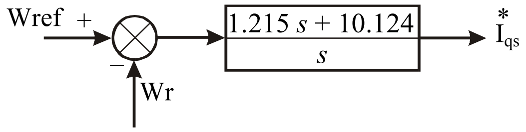

Fig. 3. Angular Speed Controller

The motor was worked with no load. The DC bus was limited in 60V/2A, and the sampling time used is 555 μs.

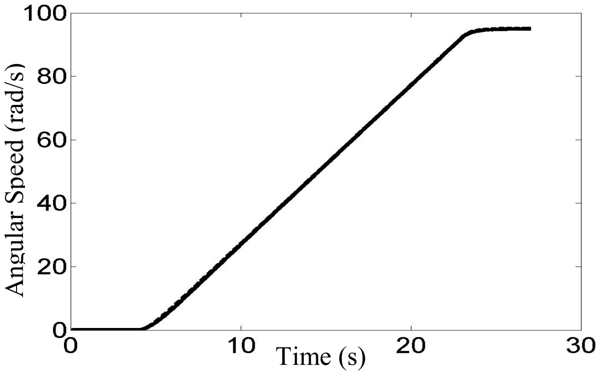

Fig. 4. Angular Speed Motor

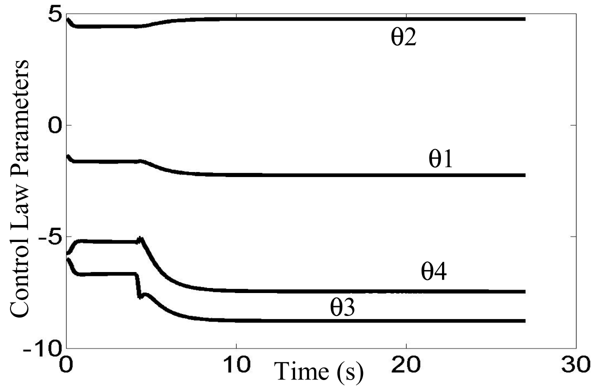

Fig. 5. Direct control law parameters

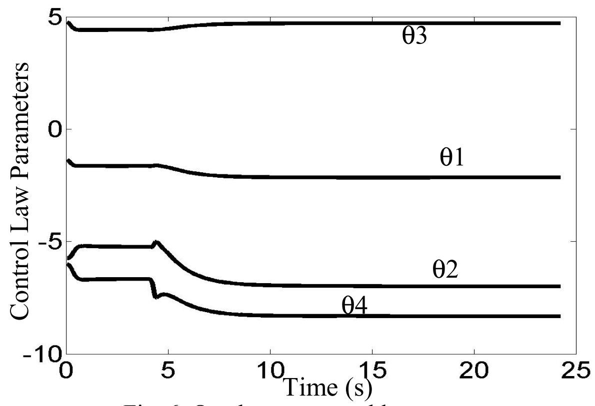

Fig. 6. Quadrature control law parameters

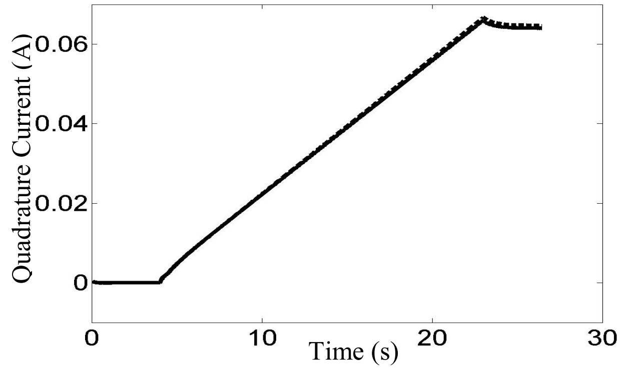

Fig. 7. Quadrature control law output (solid line) and quadrature model reference output (dashed line)

Fig. 4 shows the angular speed and the reference speed. The output of the mechanical plant was obtained by a Dynapar encoder (12 bits) and a Luenberger observer [12].

In the Fig.5 and Fig. 6 are shown the direct and quadrature control law parameters. The initial RMRAC controllers parameters were obtained by simulations, using the induction motor parameters presented on the Table 1. They are adjusted in the adaptive loop of the RMRAC controllers compensating modeling errors and unmodeled dynamics.

The quadrature control law output (U_{qs}) and output quadrature model reference output are presented in the Fig.7. The difference between U_{qs} and I_{qs} is used to obtain the quadrature voltage (V_{qs}) by (23). V_{qs} and V_{ds} were used to supply the motor.

The experimental results show the effectiveness of the proposed scheme to control the speed servo with three-phase induction motor. It can be observed that in spite of the machine is working on the nominal speed, the coupling between *dq* current dynamics is compensated by the proposed controller.

VI. CONCLUSION

This paper presents a decentralized robust model reference adaptive control (RMRAC) scheme applied to the dynamic decoupling of the stator currents of a carrier-based modulated induction motor. For the independent compensation of the d and q-axis currents it had been applied two RMRAC SISO controllers. These controllers ensure robustness and performance even in the presence of dynamics coupling between them. The experimental results demonstrate the compensation of the coupling between *d* and *q*-axis stator currents at nominal speed. Furthermore, it has been demonstrated that this scheme can be designed for a reduced order plant without a priori knowledge of exact model of the plant.

ACKNOWLEDGEMENT

The authors would like to thanks Fapergs, Capes and CNPq for the financial support.

REFERENCES

[1] R.Krishnan and A.S.Bharadwaj, "A Review of Parameter Sensitivity and Adaptation in Indirect Vector Controlled Induction Motor Drive Systems", *in Trans. on Power Electron.*, vol. 6, no.4, pp.695-703, 1991.

[2] S.M.Yang, C.H.Lee "A Deadbeat Current Controller for Field Oriented Induction Motor Drives", *IEEE Trans. Power. Electron.*, vol. 17, no.5, pp.772-778, 2002.

[3] J. Jung and K. Nam, "A Dynamic Decoupling Control Scheme for High-Speed Operation of Induction Motors", *IEEE Trans. Ind. Electronics*, vol. 46, no.1, pp.100-110, 1999.

[4] C.C.Chan, W.S.Leung and C.G.NG, "Adaptive Decoupling Control of Induction Motor Drives" *in IEEE Trans. on Ind. Electronics*, vol. 37, no.1, pp.41-47, 1990.

[5] B.D.O.Anderson, "Adaptive Systems, Lack of Persistency of Excitation and Bursting Phenomena", *IFAC Automatica*, vol.21, no.3, pp.247-258, 1985.

[6] C.E.Rohrs, L.S.Valavani, M.Athans and G.Stein, "Robustness of Continuous-Time Adaptive Control Algorithms in the Presence of Unmodeled Dynamics", *in IEEE Trans. on Automatic Control*, vol.AC-30, no.9, pp.881-889, 1985.

[7] P.A.Ioannou and J.Sun, *Robust Adaptive Control*, Prentice-Hall, Englewood Cliffs, USA, 1996.

[8] P.A.Ioannou and K.Tsakalis, "A Robust Direct Adaptive Controller", *IEEE Trans. on Automatic Control*, vol. AC-31, no.11, pp.1033-1043, 1986.

[9] E. G. Carati, C. M. Richter and H. A. Gründling, "A Robust High Performance Three-Phase Uninterruptible Power Supply," *in IEEE CCA*, pp. 896-901, Anchorage, USA, 2000.

[10] C. M. Richter and H. A. Gründling, "A Multivariable Robust Model Refrence Adaptive Controller Using a Modified Least-Squares Algorithm" in *IEEE ACC*, 2003.

[11] H.T.Câmara, E.G.Carati, H.L.Hey, H.Pinheiro, J.R.Pinheiro and H.A.Gründling, "Speed and Position Servo for Induction Motor using Robust Model Reference Adaptive Control," *in IEEE IECON*, Sevilha, Spain, 2002.

[12] M. Campos, E. G. Carati, H.A.Gründling, "Design of a position servo with induction motor using self-tuning regulator and Kalman filter," *IAS Conf. Record*, v.3, pp. 1613-1618, 2000.