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A NEW GLOBAL CONTROL SCHEME FOR SENSORLESS CURRENT-FED INDUCTION MOTORS

R. Marino, P. Tomei and C.M. Verrelli

Dip. Ing. Elettronica, Univ. Roma 'Tor Vergata', Italy, e-mail: tomei@eln.uniroma2.it

Abstract: The problem of controlling a sensorless induction motor (i.e. without rotor speed measurements) is addressed. Smooth reference signals for rotor speed and rotor flux modulus are required to be tracked exponentially and globally. Only semiglobal solutions have been recently obtained in the literature. A global solution is presented for current-fed induction motors which makes use of a novel rotor speed observer and can be naturally extended to the full-order model. *Copyright © 2002 IFAC*

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1. INTRODUCTION

In recent years significant advances (Blaschke, 1972; Leonhard, 1985; Marino, *et al*, 1993; Hu, and Dawson, 1996; Ortega, *et al*, 1996; Marino, *et al*, 1999; Vedagarbha, *et al*, 1999) have been made on the control of induction motors on the basis of rotor speed and stator current measurements only (no rotor flux measurements). For instance in (Marino, *et al*, 1999) it is shown that smooth reference signals for rotor speed and rotor flux modulus (so that motor efficiency can be improved) can be asymptotically tracked from any initial condition and for unknown constant values of rotor resistance and torque load. Even though induction motors are difficult to control (multivariable nonlinear dynamics with two inputs and two outputs, outputs to be controlled which are not measured, critical parameters which vary during operations), it is by now clear that high performances can be achieved.

However in many cases velocity sensors are not available to reduce costs and to increase reliability: this further complicates the control task. This problem is usually called sensorless control since rotor flux and speed are not measured and only stator currents are available for measurements. There is a large number of papers devoted to this problem (Tajima, and Hori, 1993; Hurst, *et al*, 1998; Jadot, *et al*, 2001, and references therein). Torque control even at low speed is achieved in (Hurst, *et al*, 1998). A speed estimation method is proposed in (Tajima, and Hori, 1993) and incorporated in a field orientation control scheme. An industrial solution based on advanced nonlinear control techniques was recently reported in (Jadot, *et al*, 2001). However, no closed loop stability analysis is presented in those papers, with the exception of (Feemster, *et al*, 2001), in which semiglobal exponential rotor velocity/rotor flux tracking is proved for the full order nonlinear model of an induction motor.

In this paper a new control scheme for sensorless current-fed induction motors is proposed: it is based on a rotor speed observer which includes a feedback term proportional to the quadrature component of the rotor flux vector in a suitable rotating frame. The contribution of the paper is to show that the proposed control scheme guarantees local exponential stability and global asymptotic stability of the closed loop current-fed induction motor dynamics provided that rotor flux measurements are available. The control scheme is extended to the full-order model. It is then argued, as in (Tajima, and Hori, 1993; Hu, and Wu, 1998; Hurst, *et al*, 1998; Vedagarbha, *et al*, 1999; Feemster, *et al*, 2001; Pinto, *et al*, 2001), that, since rotor flux measurements are not available, rotor flux can be computed on line by integrating stator flux dynamic equations from zero initial conditions. The closed loop system performances are illustrated by simulation.

2. PROBLEM STATEMENT

Assuming linear magnetic circuits, a balanced non-saturated induction motor with one pole pair is modeled in a fixed reference frame attached to the stator by the well known fifth-order model, see for istance (Marino, *et al*, 1993) for its derivation and (Leonhard, 1985; Krause, 1986) for modeling assumptions,

$$\begin{aligned}
\frac{d\omega}{dt} &= \mu(\psi_{ra} i_{sb} - \psi_{rb} i_{sa}) - \frac{T_L}{J} \\
\frac{d\psi_{ra}}{dt} &= -\alpha\psi_{ra} - \omega\psi_{rb} + \alpha M i_{sa} \\
\frac{d\psi_{rb}}{dt} &= -\alpha\psi_{rb} + \omega\psi_{ra} + \alpha M i_{sb} \\
\frac{di_{sa}}{dt} &= -\frac{R_s}{\sigma} i_{sa} + \frac{1}{\sigma} u_{sa} + \beta\alpha\psi_{ra} + \beta\omega\psi_{rb} \\
&\quad -\beta\alpha M i_{sa} \\
\frac{di_{sb}}{dt} &= -\frac{R_s}{\sigma} i_{sb} + \frac{1}{\sigma} u_{sb} + \beta\alpha\psi_{rb} - \beta\omega\psi_{ra} \\
&\quad -\beta\alpha M i_{sb}
\end{aligned} \tag{1}$$

in which: ω is the rotor speed, (ψ_{ra}, ψ_{rb}) are the rotor fluxes, (i_{sa}, i_{sb}) are the stator currents, (u_{sa}, u_{sb}) are the stator voltages which constitute the control inputs; the outputs to be controlled are the rotor speed ω and the rotor flux modulus $\sqrt{\psi_{ra}^2 + \psi_{rb}^2}$. The only measured variables are the stator currents (i_{sa}, i_{sb}), which may be viewed as the control inputs when high gain feedback are used to drive the stator currents (i_{sa}, i_{sb}) to their desired reference values (i_{sa}^*, i_{sb}^*): in this case the motor is called current-fed and, neglecting the stator currents dynamics, is described by the first three equations in (1). The control problem is called sensorless since the three state variables $(\omega, \psi_{ra}, \psi_{rb})$ are not measured while only stator currents (i_{sa}, i_{sb}) measurements are available (in addition to stator voltages) for the full order model (1). The motor parameters are: torque load T_L, rotor moment of inertia J, rotor and stator windings resistances (R_r, R_s) and inductances (L_r, L_s), mutual inductance M. To simplify notations we use the reparametrization: $\alpha = \frac{R_r}{L_r}$, $\beta = \frac{M}{\sigma L_r}$, $\mu = \frac{M}{JL_r}$, $\sigma = L_s(1 - \frac{M^2}{L_s Lr})$. As in (Marino ,*et al*,1999) we introduce an angle $\varepsilon_0(t)$, whose dynamics $\dot{\varepsilon}_0(t) = \omega_0(t)$ will be later defined ($\varepsilon_0(0)$ is an arbitrary initial condition). We also introduce the variables $(\psi_{rd}, \psi_{rq})^T$, $(i_{sd}, i_{sq})^T$, $(u_{sd}, u_{sq})^T$, which are obtained from the corresponding (a, b) variables $(\psi_{ra}, \psi_{rb})^T$, $(i_{sa}, i_{sb})^T$, $(u_{sa}, u_{sb})^T$ multiplying by the matrix $\begin{pmatrix} \cos\varepsilon_0 & \sin\varepsilon_0 \\ -\sin\varepsilon_0 & \cos\varepsilon_0 \end{pmatrix}$. They represent the components of rotor flux, stator current and stator voltages vectors, respectively, with respect to a time-varying (d, q) frame rotating at speed ω_0 and identified by the angle $\varepsilon_0(t)$. In the new state coordinates $(\omega, \psi_{rd}, \psi_{rq}, i_{sd}, i_{sq})$ and new control coordinates (u_{sd}, u_{sq}) the motor dynamics (1) become, see for istance (Krause, 1986) for induction machine models in arbitrarily rotating frames,

$$\begin{aligned}
\frac{d\omega}{dt} &= \mu(\psi_{rd} i_{sq} - \psi_{rq} i_{sd}) - \frac{T_L}{J} \\
\frac{d\psi_{rd}}{dt} &= -\alpha\psi_{rd} + (\omega_0 - \omega)\psi_{rq} + \alpha M i_{sd} \\
\frac{d\psi_{rq}}{dt} &= -\alpha\psi_{rq} - (\omega_0 - \omega)\psi_{rd} + \alpha M i_{sq} \\
\frac{di_{sd}}{dt} &= -\frac{R_s}{\sigma} i_{sd} + \frac{1}{\sigma} u_{sd} + \omega_0 i_{sq} + \beta\alpha\psi_{rd} \\
&\quad +\beta\omega\psi_{rq} - \beta\alpha M i_{sd} \\
\frac{di_{sq}}{dt} &= -\frac{R_s}{\sigma} i_{sq} + \frac{1}{\sigma} u_{sq} - \omega_0 i_{sd} + \beta\alpha\psi_{rq} \\
&\quad -\beta\omega\psi_{rd} - \beta\alpha M i_{sq} \\
\frac{d\varepsilon_0}{dt} &= \omega_0 \ .
\end{aligned} \tag{2}$$

Let us denote by $\omega^*(t)$ and $\psi^*(t) > 0$ the smooth bounded reference signals for the output variables to be controlled, which are the speed ω and the rotor flux modulus $\sqrt{\psi_{ra}^2 + \psi_{rb}^2} = \sqrt{\psi_{rd}^2 + \psi_{rq}^2}$ respectively, and by $\tilde{\omega}(t) = \omega(t) - \omega^*(t)$, $\tilde{\psi}_{rd}(t) = \psi_{rd}(t) - \psi^*(t)$, $\tilde{\psi}_{rq}(t) = \psi_{rq}(t)$,the tracking errors of the speed and the (d, q) flux components. Following the field-oriented control strategy (Blaschke, 1972), our goal is to design a dynamic output feedback compensator [recall that the measured outputs are (i_{sa}, i_{sb})]

$$\begin{aligned}
\frac{d\varepsilon_0(t)}{dt} &= \omega_0(t) \\
\begin{pmatrix} u_{sa} \\ u_{sb} \end{pmatrix} &= \begin{pmatrix} \cos\varepsilon_0 & -\sin\varepsilon_0 \\ \sin\varepsilon_0 & \cos\varepsilon_0 \end{pmatrix} \begin{pmatrix} u_{sd} \\ u_{sq} \end{pmatrix}
\end{aligned} \tag{3}$$

by choosing $(\omega_0, u_{sd}, u_{sq})$ so that for any initial condition $(\omega(0), \psi_{ra}(0), \psi_{rb}(0), i_{sa}(0), i_{sb}(0), \varepsilon_0(0))$ we have

$$\lim_{t\to\infty} [\omega(t) - \omega^*(t)] = 0 \tag{4}$$

and

$$\lim_{t\to\infty} [\psi_{rd}(t) - \psi^*(t)] = 0 \tag{5}$$

$$\lim_{t\to\infty} \psi_{rq} = 0 \tag{6}$$

which imply that

$$\lim_{t\to\infty}[\sqrt{\psi_{ra}^2+\psi_{rb}^2}-\psi^*]=0 \quad . \tag{7}$$

Note that (5), (6) imply that the flux vector (ψ_{ra}, ψ_{rb}) asymptotically rotates at speed ω_0, i.e., field orientation is achieved. In other words, the (d, q) frame rotating at speed ω_0 tends to have the d-axis coincident with the rotating flux vector as t goes to infinity.

3. OPEN LOOP CONTROL

If we replace the state variables $(\omega, \psi_{rd}, \psi_{rq})$ by the corresponding reference signals $(\omega^*, \psi^*, 0)$ in the first three equations in (2), we obtain

$$\begin{aligned}
\frac{d\omega^*}{dt} &= \mu\psi^* i_{sq}^* - \frac{T_L}{J} \\
\frac{d\psi^*}{dt} &= -\alpha\psi^* + \alpha M i_{sd}^* \\
0 &= -(\omega_0-\omega^*)\psi^* + \alpha M i_{sq}^*
\end{aligned} \tag{8}$$

from which we solve for stator currents reference signals (i_{sd}^*, i_{sq}^*)

$$i_{sd}^* = \frac{\psi^*}{M} + \frac{\dot{\psi}^*}{\alpha M} \tag{9}$$

$$i_{sq}^* = \frac{1}{\mu\psi^*}(\frac{T_L}{J} + \dot{\omega}^*) \tag{10}$$

and obtain the dynamics of the rotating angle $\varepsilon_0(t)$

$$\frac{d\varepsilon_0}{dt} = \omega_0 = \omega^* + \frac{\alpha M i_{sq}^*}{\psi^*} \tag{11}$$

where $\omega_s = \alpha M i_{sq}^*/\psi^*$ is usually called the slip speed. Equations (9), (10), (11) lead to the open loop reference signals (i_{sa}^*, i_{sb}^*) for the stator currents (i_{sa}, i_{sb}):

$$\begin{pmatrix} i_{sa}^* \\ i_{sb}^* \end{pmatrix} = \begin{pmatrix} \cos\varepsilon_0 & -\sin\varepsilon_0 \\ \sin\varepsilon_0 & \cos\varepsilon_0 \end{pmatrix} \begin{pmatrix} i_{sd}^* \\ i_{sq}^* \end{pmatrix}$$

which, substituted in the first three equations in (2) with initial conditions $\omega(0) = \omega^*(0)$, $\psi_{rd}(0) = \psi^*(0)$, $\psi_{rq}(0) = 0$ guarantee $\tilde{\omega}(t) = 0$, $\tilde{\psi}_{rd}(t) = 0$, $\tilde{\psi}_{rq}(t) = 0$, $\forall t \geq 0$. The open loop control error dynamics for the current-fed induction motor model are

$$\begin{aligned}
\dot{\tilde{\omega}} &= \mu(\tilde{\psi}_{rd} i_{sq}^* - \tilde{\psi}_{rq} i_{sd}^*) \\
\dot{\tilde{\psi}}_{rd} &= -\alpha\tilde{\psi}_{rd} + (\omega_0-\omega)\tilde{\psi}_{rq} \\
\dot{\tilde{\psi}}_{rq} &= -\alpha\tilde{\psi}_{rq} - (\omega_0-\omega)\tilde{\psi}_{rd} + \tilde{\omega}\psi^* \ .
\end{aligned} \tag{12}$$

If ω^*=const, ψ^*=const and $T_L \neq 0$ then there is an additional equilibrium point for (12) besides the origin and the stability properties are rather difficult to be analyzed.

4. A NEW GLOBAL CONTROL SCHEME

If the rotor speed measurement is available, the indirect field oriented control

$$\begin{aligned}
i_{sd}^* &= \frac{\psi^*}{M} + \frac{\dot{\psi}^*}{\alpha M} \\
i_{sq}^* &= \frac{1}{\mu\psi^*}(-k_\omega(\omega-\omega^*) + \frac{T_L}{J} + \dot{\omega}^*) \\
\omega_0 &= \omega + \frac{\alpha M i_{sq}^*}{\psi^*}
\end{aligned} \tag{13}$$

yields the closed loop error dynamics

$$\begin{aligned}
\dot{\tilde{\omega}} &= -k_\omega\tilde{\omega} + \mu(\tilde{\psi}_{rd} i_{sq}^* - \tilde{\psi}_{rq} i_{sd}^*) \\
\dot{\tilde{\psi}}_{rd} &= -\alpha\tilde{\psi}_{rd} + (\omega_0-\omega)\tilde{\psi}_{rq} \\
\dot{\tilde{\psi}}_{rq} &= -\alpha\tilde{\psi}_{rq} - (\omega_0-\omega)\tilde{\psi}_{rd}
\end{aligned} \tag{14}$$

which are globally asymptotically stable, see (Ortega, *et al*, 1996).

We now propose a new control scheme which modifies (13) by replacing ω by $\hat{\omega}$ and introducing a saturation function $sat(x)$ (a C^∞ odd function whose time derivative is always positive and has a finite limit as $|x|$ goes to infinity)

$$\begin{aligned}
i_{sd}^* &= \frac{\psi^*}{M} + \frac{\dot{\psi}^*}{\alpha M} \\
i_{sq}^* &= \frac{1}{\mu\psi^*}(-k_\omega sat(\hat{\omega}-\omega^*) + \frac{T_L}{J} + \dot{\omega}^*) \\
\frac{d\varepsilon_0}{dt} &= \omega_0 = \hat{\omega} + \frac{\alpha M i_{sq}^*}{\psi^*} \\
\begin{pmatrix} i_{sa}^* \\ i_{sb}^* \end{pmatrix} &= \begin{pmatrix} \cos\varepsilon_0 & -\sin\varepsilon_0 \\ \sin\varepsilon_0 & \cos\varepsilon_0 \end{pmatrix} \begin{pmatrix} i_{sd}^* \\ i_{sq}^* \end{pmatrix}
\end{aligned} \tag{15}$$

where $\hat{\omega}$ is the speed estimate provided by the speed observer

$$\dot{\hat{\omega}} = \mu(\psi_{rd} i_{sq}^* - \psi_{rq} i_{sd}^*) - \frac{T_L}{J} + v \tag{16}$$

in which the term v will be designed using adaptive control techniques. Equations (15) and (16) yield the closed loop error dynamics ($e_\omega(t) = \hat{\omega}(t) - \omega(t)$)

$$\begin{aligned}
\dot{\tilde{\omega}} &= -k_\omega sat(\tilde{\omega}+e_\omega) + \mu(\tilde{\psi}_{rd} i_{sq}^* - \tilde{\psi}_{rq} i_{sd}^*) \\
\dot{\tilde{\psi}}_{rd} &= -\alpha\tilde{\psi}_{rd} + (\omega_0-\omega)\tilde{\psi}_{rq} \\
\dot{\tilde{\psi}}_{rq} &= -\alpha\tilde{\psi}_{rq} - (\omega_0-\omega)\tilde{\psi}_{rd} - e_\omega\psi^* \\
\dot{e}_\omega &= v \ .
\end{aligned} \tag{17}$$

Consider the positive definite function ($\gamma_1 > 0$),

$$V(t) = \frac{1}{2}(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) + \frac{1}{2}\gamma_1 e_\omega^2 \tag{18}$$

whose time derivative is

$$\dot{V}(t) = -\alpha(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) + (\gamma_1 v - \psi^* \tilde{\psi}_{rq})e_\omega \quad (19)$$

Choosing

$$v = \frac{\psi^* \tilde{\psi}_{rq}}{\gamma_1} \quad (20)$$

we obtain

$$\dot{V}(t) = -\alpha(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) \quad . \quad (21)$$

From (18) and (21), it follows that $(\tilde{\psi}_{rd}, \tilde{\psi}_{rq}, e_\omega)$ are bounded for every $t \geq 0$. The closed loop system is

$$\begin{aligned}
\dot{\tilde{\omega}} &= -k_\omega sat(\tilde{\omega} + e_\omega) + \mu(\tilde{\psi}_{rd} i_{sq}^* - \tilde{\psi}_{rq} i_{sd}^*) \\
\dot{\tilde{\psi}}_{rd} &= -\alpha\tilde{\psi}_{rd} + (\omega_0 - \omega)\tilde{\psi}_{rq} \\
\dot{\tilde{\psi}}_{rq} &= -\alpha\tilde{\psi}_{rq} - (\omega_0 - \omega)\tilde{\psi}_{rd} - e_\omega \psi^* \quad (22) \\
\dot{e}_\omega &= \frac{\psi^* \tilde{\psi}_{rq}}{\gamma_1} \quad .
\end{aligned}$$

The last three equations in (22) may be rewritten as

$$\begin{aligned}
\dot{x} &= A(t)x + \Gamma^T(t)y \\
\dot{y} &= -\frac{1}{\gamma_1}\Gamma(t)x \quad (23)
\end{aligned}$$

with $x = [\tilde{\psi}_{rd}, \tilde{\psi}_{rq}]^T$, $y = e_\omega$ and

$$A(t) = \begin{pmatrix} -\alpha & (\omega_0 - \omega) \\ -(\omega_0 - \omega) & -\alpha \end{pmatrix}, \ \Gamma^T(t) = \begin{pmatrix} 0 \\ -\psi^* \end{pmatrix}.$$

The radially unbounded function (18) may be rewritten as

$$V = \frac{1}{2}(x^T x + \gamma_1 y^2)$$

whose time derivative is $\dot{V} = x^T(A^T(t) + A(t))x/2 = -\alpha x^T x$. Using the same arguments adopted in the proof of the persistency of excitation Lemma 1 in (Marino, *et al*, 2001), (recall from (15) that $\omega_0 - \omega = e_\omega + \frac{\alpha M i_{sq}^*}{\psi^*}$ is bounded), we can establish that since $\psi^* > 0$, $\forall t \geq 0$, persistency of excitation conditions are satisfied, i.e. there exist two positive constants T and c such that

$$\int_t^{T+t} \Gamma(\tau)\Gamma^T(\tau)d\tau = \int_t^{T+t} (\psi^*(\tau))^2 d\tau \geq c \quad \forall t \geq 0$$

it follows that the equilibrium point $(x, y) = 0$ of system (23) is globally exponentially stable. Recalling the first equation in (22), now we can establish that, since (i_{sd}^*, i_{sq}^*) are bounded in (15) and $(\tilde{\psi}_{rd}, \tilde{\psi}_{rq}, e_\omega)$ tend exponentially to zero, $\tilde{\omega}$ tends asymptotically to zero and the origin is a locally exponentially equilibrium point for system (22). The control algorithm (15),(16),(20) requires rotor flux measurements (ψ_{rd}, ψ_{rq}) in the speed observer (16), (20), which are usually not available. However, rotor fluxes can be obtained by open-loop on line integration from zero initial conditions. The equations to be integrated on line are

$$\begin{aligned}
\dot{\psi}_{sd} &= u_{sd} - R_s i_{sd} + \omega_0 \psi_{sq}, \quad \psi_{sd}(0) = 0 \\
\dot{\psi}_{sq} &= u_{sq} - R_s i_{sq} - \omega_0 \psi_{sd}, \quad \psi_{sq}(0) = 0 \\
\psi_{rd} &= M i_{sd} - \frac{L_r L_s}{M} i_{sd} + \frac{L_r}{M}\psi_{sd} \\
\psi_{rq} &= M i_{sq} - \frac{L_r L_s}{M} i_{sq} + \frac{L_r}{M}\psi_{sq} \quad (24)
\end{aligned}$$

where (ψ_{sd}, ψ_{sq}) denote the stator flux vector rotating at speed ω_0.

5. EXTENSION TO THE FULL-ORDER MODEL

The control scheme proposed for current-fed induction motors, can be extended to the full-order model (2). We define the reference current signals

$$\begin{aligned}
i_{sd}^* &= \frac{\psi^*}{M} + \frac{\dot{\psi}^*}{\alpha M} \\
i_{sq}^* &= \frac{1}{\mu\psi^*}(-k_\omega sat(\hat{\omega} - \omega^*) + \frac{T_L}{J} + \dot{\omega}^*) \quad (25)
\end{aligned}$$

for i_{sd} and i_{sq}, and the speed of the rotating (d, q) frame

$$\frac{d\varepsilon_0}{dt} = \omega_0 = \hat{\omega} + \frac{\alpha M i_{sq}}{\psi^*} \quad (26)$$

where $\hat{\omega}$ is the speed estimate provided by the speed observer

$$\dot{\hat{\omega}} = \mu(\psi_{rd} i_{sq} - \psi_{rq} i_{sd}) - \frac{T_L}{J} + w \quad (27)$$

in which the term w will be designed using adaptive control techniques. Introducing the current tracking errors

$$\begin{aligned}
e_d &= i_{sd} - i_{sd}^* \\
e_q &= i_{sq} - i_{sq}^* \quad (28)
\end{aligned}$$

from (2), (26)-(28) we obtain

$$\begin{aligned}
\dot{\tilde{\omega}} &= -k_\omega sat(\tilde{\omega} + e_\omega) + \mu(\tilde{\psi}_{rd} i_{sq} - \tilde{\psi}_{rq} i_{sd}) \\
&\quad + \mu\psi^* e_q \\
\dot{\tilde{\psi}}_{rd} &= -\alpha\tilde{\psi}_{rd} + (\omega_0 - \omega)\tilde{\psi}_{rq} + \alpha M e_d \\
\dot{\tilde{\psi}}_{rq} &= -\alpha\tilde{\psi}_{rq} - (\omega_0 - \omega)\tilde{\psi}_{rd} - e_\omega \psi^* \quad (29) \\
\dot{e}_\omega &= w
\end{aligned}$$

$$
\begin{aligned}
\dot{e}_d &= \frac{1}{\sigma}u_{sd} + \phi_{d0} + \phi_{d1}e_\omega \\
\dot{e}_q &= \frac{1}{\sigma}u_{sq} + \phi_{q0} + \phi_{q1}e_\omega
\end{aligned}
$$

with the known functions

$$
\begin{aligned}
\phi_{d0} &= -\frac{R_s}{\sigma}i_{sd} + \omega_0 i_{sq} + \alpha\beta\psi_{rd} + \beta\hat{\omega}\psi_{rq} \\
&\qquad -\alpha M\beta i_{sd} - \frac{\dot{\psi}^*}{M} - \frac{\ddot{\psi}^*}{\alpha M} \\
\phi_{d1} &= -\beta\psi_{rq} \\
\phi_{q0} &= -\frac{R_s}{\sigma}i_{sq} - \omega_0 i_{sd} + \alpha\beta\psi_{rq} - \beta\hat{\omega}\psi_{rd} - \alpha M\beta i_{sq} \\
&\qquad + \frac{\dot{\psi}^*}{\mu{\psi^*}^2}(-k_\omega sat(\hat{\omega} - \omega^*) + \frac{T_L}{J} + \dot{\omega}^*) \\
&\qquad - \frac{1}{\mu\psi^*}(-k_\omega \frac{d[sat(\hat{\omega} - \omega^*)]}{dt} + \frac{\dot{T}_L}{J} + \ddot{\omega}^*) \\
\phi_{q1} &= \beta\psi_{rd} \ .
\end{aligned}
$$

In order to determine w and the feedback control inputs (u_{sd}, u_{sq}), we consider the function

$$
V_0(t) = \frac{1}{2}(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) + \frac{1}{2}\gamma_2 e_\omega^2 + \frac{1}{2}(e_d^2 + e_q^2) \tag{30}
$$

(γ_2 is a positive real parameter), whose time derivative is

$$
\begin{aligned}
\dot{V}_0(t) &= -\alpha(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) \\
&+ (\gamma_2 w - \psi^*\tilde{\psi}_{rq} + \phi_{d1}e_d + \phi_{q1}e_q)e_\omega \\
&+ (\frac{1}{\sigma}u_{sd} + \phi_{d0} + \alpha M\tilde{\psi}_{rd})e_d + (\frac{1}{\sigma}u_{sq} + \phi_{q0})e_q
\end{aligned} \tag{31}
$$

We define

$$
\begin{aligned}
w &= \frac{\psi^*\tilde{\psi}_{rq}}{\gamma_2} - \frac{\phi_{d1}e_d}{\gamma_2} - \frac{\phi_{q1}e_q}{\gamma_2} \\
u_{sd} &= \sigma(-\phi_{d0} - \alpha M\tilde{\psi}_{rd} - k_e e_d) \\
u_{sq} &= \sigma(-\phi_{q0} - k_e e_q)
\end{aligned} \tag{32}
$$

so that (32) becomes

$$
\dot{V}_0(t) = -\alpha(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) - k_e(e_d^2 + e_q^2) \ . \tag{33}
$$

From (31) and (34), it follows that $(\tilde{\psi}_{rd}, \tilde{\psi}_{rq}, e_\omega, e_d, e_q)$ are bounded. Their bounds depend on the initial errors. From (30) and (33), the closed loop system is

$$
\begin{aligned}
\dot{\tilde{\omega}} &= -k_\omega sat(\tilde{\omega} + e_\omega) + \mu(\tilde{\psi}_{rd}i_{sq} - \tilde{\psi}_{rq}i_{sd}) \\
&\quad + \mu\psi^* e_q \\
\dot{\tilde{\psi}}_{rd} &= -\alpha\tilde{\psi}_{rd} + (\omega_0 - \omega)\tilde{\psi}_{rq} + \alpha M e_d \\
\dot{\tilde{\psi}}_{rq} &= -\alpha\tilde{\psi}_{rq} - (\omega_0 - \omega)\tilde{\psi}_{rd} - e_\omega\psi^* \\
\dot{e}_\omega &= \frac{\psi^*\tilde{\psi}_{rq}}{\gamma_2} - \frac{\phi_{d1}e_d}{\gamma_2} - \frac{\phi_{q1}e_q}{\gamma_2} \\
\dot{e}_d &= -\alpha M\tilde{\psi}_{rd} - k_e e_d - \beta\psi_{rq}e_\omega \\
\dot{e}_q &= -k_e e_q + \beta\psi_{rd}e_\omega \ .
\end{aligned} \tag{34}
$$

It is not difficult to show that the origin is an equilibrium point for system (34) which is globally asymptotically stable and locally exponentially stable. In fact, recalling the first equation in (34), we can establish that, since (i_{sd}, i_{sq}, ψ^*) are bounded and $(\tilde{\psi}_{rd}, \tilde{\psi}_{rq}, e_\omega, e_q)$ tend, due to persistency of excitation, exponentially to zero, $\tilde{\omega}$ tends asymptotically to zero from any initial condition and is bounded by a decreasing exponential from sufficiently small initial conditions.

6. SIMULATION RESULTS

We tested the proposed controller by simulations with the control parameters (all values are in SI units) $k_\omega = 800, \gamma_1 = 0.0008$ for a three-phase single pole pair 0.6 kW induction motor, whose parameters are: $J = 0.0075$ Kg m^2, $R_s = 5.3$, $R_r = 3.3$, $L_s = 0.365$ H, $L_r = 0.375$ H, $M = 0.34$ H. All initial conditions of the motor and of the controller are set to zero. The references for speed and flux modulus along with the applied torque are reported in Fig. 1. The flux reference starts from 0.001 Wb at $t = 0$ and grows up to the rated constant value 1.16 Wb; field weakening starts at $t = 1.5$ s. The speed reference is zero until $t = 0.32$ s and grows up to the constant value 100 rad/s; at $t = 1.5$ s the speed is required to reach the value 200 rad/s, while the reference for the flux is reduced to 0.5 Wb. A constant load torque (5.8 Nm, the rated value) is applied at $t = 0.5$ s and reduced to 0.5 Nm at $t = 1$ s. Fig. 2 shows the time histories of speed and flux modulus tracking errors and of the speed estimation error.

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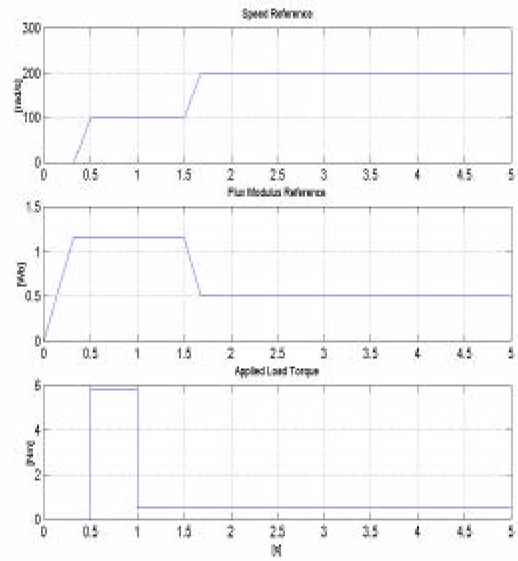

Fig. 1. Speed and flux modulus reference signals and applied load torque.

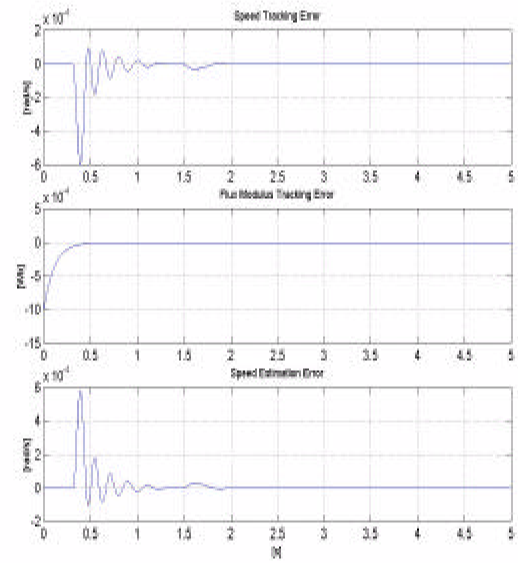

Fig.2. Tracking and estimation errors.