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Speed Regulation of Induction Motors: An Adaptive Sensorless Sliding Mode Control Scheme

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Abstract— **A current-based sensorless sliding mode control for induction motors is presented in this paper. It guarantees asymptotic tracking of prespecified speed and square of the rotor flux magnitude references without mechanical sensors. The problem of chattering, typical of sliding mode controllers, is overcome since the derivative of the stator currents are used as discontinuous forcing actions, while the actual control signals are continuous, thus limiting the mechanical stress. The proposed adaptive sliding mode speed observer is based on a different approach with respect to the widely used equivalent control techniques.**

I. Introduction

In the last years, sensorless control of motor drives has become particularly attractive. Indeed, advantages presented by the induction motor with respect to the other kind of electric machines are unquestionable [1]: thanks to its simple squirrel cage rotor structure, without permanent magnets or brushes, it results more reliable and less expensive, and suitable for applications in hostile environments, because no sparks are produced during operations. These advantages become definitely appreciable if sensors for mechanical variables are not required.

Unfortunately, speed (torque) and flux controlling in induction motors drives is a difficult task because of the nonlinearities and the strict coupling between the state variables. Generally, no flux sensors are provided, and even when a speed measurement is available, flux observers convergence risks to be compromised by significant parameters values variations (the most critical, the rotor resistance, may change up to 200% of the nominal value), while the measurements of stator currents turn out to be affected by noise, due to electro-magnetic disturbances or to harmonics. High performances and high robustness properties are required to the control and the observer algorithms.

A great number of valid alternatives to the classical and widely used Field Oriented Control strategy have been proposed: a typical Nonlinear Output Feedback Control scheme for current-fed induction motors is presented in [2], while a Sliding Mode control scheme has been described by Utkin in [3]. A global speed control scheme without mechanical sensors has been recently proposed in [4]. A lot of efforts have been also dedicated to the problem of parameters estimation: in particular, solutions for rotor resistance on-line tuning are described in [5], [6], [7]; in [8] stator resistance tuning is also taken into account.

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The Sliding Mode control design techniques, capable of guaranteeing high levels of robustness against matched disturbances and parameters variation, seem to be well applicable to the problems of sensorless speed and torque control and robust flux estimation of induction motors: in [9] the speed estimation is obtained by filtering a discontinuous signal, relaying on the concept of equivalent control; a different scheme, based on the same theoretical concept, is proposed in [10]; robust speed and torque estimation by high order sliding modes is described in [11].

In this paper, a new Adaptive Sensorless Sliding Mode Control strategy is presented. A classical current-fed induction motor control scheme, where the stator currents are assumed as control signals, is adopted, maintaining the conventional control loops and simply replacing the control algorithm. To avoid the problem of chattering, the time derivatives of the currents have been regarded as auxiliary control signals, while the actual control signals are continuous.

The key element of the proposed method is a novel sliding mode speed and flux adaptive observer. Relying on a double sliding mode current observer, convergence of flux, speed and rotor time constant estimates to real values is guaranteed. Simulations show that the observer-based control algorithm provides high regulation accuracy and appreciable robustness.

II. Model of the Induction Motor and Problem Formulation

In a fixed reference frame $a-b$, the fifth order induction motor model is defined by the following equations

$$\begin{cases} \frac{d\psi_a}{dt} &= -\alpha\psi_a - \omega\psi_b + M\alpha i_a \\ \frac{d\psi_b}{dt} &= -\alpha\psi_b + \omega\psi_a + M\alpha i_b \\ \frac{di_a}{dt} &= -\beta\frac{d\psi_a}{dt} + \frac{1}{\sigma L_s}(u_a - R_s i_a) \\ \frac{di_b}{dt} &= -\beta\frac{d\psi_b}{dt} + \frac{1}{\sigma L_s}(u_b - R_s i_b) \\ \frac{d\omega}{dt} &= \frac{1}{J}\frac{M}{L_r}(\psi_a i_b - \psi_b i_a) - \frac{\Gamma_l}{J} \end{cases} \quad (1)$$

where the state variables are the rotor speed ω, the rotor fluxes (ψ_a, ψ_b) and the stator currents (i_a, i_b). Stator voltages (u_a, u_b) are the control signals, Γ_l is the load torque, J is the moment of inertia, (R_r, R_s) and (L_r, L_s) are rotor and stator windings resistances and inductances, respectively, and M is the mutual inductance.

An induction motor with one pole pair is considered. To simplify notations, the following parameters have been introduced

$$\alpha = \frac{R_r}{L_r}, \quad \sigma = 1 - \frac{M^2}{L_s L_r}, \quad \beta = \frac{1}{\sigma}\frac{M}{L_s L_r}, \tag{2}$$

For current-fed induction motors with high-gain current loops the motor control algorithm can be constructed on the basis of the following reduced order motor model

$$\begin{cases} \frac{d\psi_a}{dt} & = & -\alpha\psi_a - \omega\psi_b + M\alpha i_a \\ \frac{d\psi_b}{dt} & = & -\alpha\psi_b + \omega\psi_a + M\alpha i_b \\ \frac{d\omega}{dt} & = & \frac{1}{J}\frac{M}{L_r}\left(\psi_a i_b - \psi_b i_a\right) - \frac{\Gamma_l}{J} \end{cases} \tag{3}$$

by considering the stator currents (i_a, i_b) as control inputs, the rotor fluxes (ψ_a, ψ_b) as the state variables, Γ_l as a known input, and α as an unknown parameter (depending on the rotor resistance value).

The quantities $\omega_r(t)$ and $\Psi_r^2(t)$ are the reference signals for the rotor speed and the square modulus of the rotor flux $\Psi^2 = \psi_a^2 + \psi_b^2$, respectively[1]. Then, the tracking errors $\tilde{\omega}$ and $\tilde{\Psi}$ can be defined as

$$\begin{cases} \tilde{\omega} & = & \omega - \omega_r \\ \tilde{\Psi} & = & \Psi^2 - \Psi_r^2 = \psi_a^2 + \psi_b^2 - \Psi_r^2 \end{cases} \tag{4}$$

such that their time derivatives are

$$\begin{cases} \dot{\tilde{\omega}} & = & \frac{1}{J}\frac{M}{L_r}\left(\psi_a i_b - \psi_b i_a\right) - \frac{\Gamma_l}{J} - \dot{\omega}_r \\ \dot{\tilde{\Psi}} & = & -2\alpha\left(\psi_a^2 + \psi_b^2\right) \\ & & +2M\alpha\left(\psi_a i_a + \psi_b i_b\right) - 2\Psi_r\dot{\Psi}_r \end{cases} \tag{5}$$

The problems addressed in the paper are the following:

- to design a control algorithm that guarantees that speed and flux tracking errors are driven to zero with exponential law;
- to design a speed and flux observer, to be included in the control scheme, adaptive with respect to the unknown rotor time constant (*Sensorless Control*).

III. Sliding Mode Speed and Flux Control

Hereafter, the induction motor basic sliding mode control methodology is first briefly recalled for the reader's convenience, the new current-based sliding mode control algorithm is designed and, finally, the proposed speed and flux observer is discussed.

[1]To allow for correct operation of the control algorithm, the first and second time derivatives of the speed and flux references are assumed to be bounded.

A. Preliminary Issues: Voltage-based Sliding Mode Control

Among the various sliding mode control solutions for induction motors proposed in the literature, the one presented in [3] can be regarded as the reference one. Its purpose is to directly control the inverter switching by use of three switching reference signals for the stator voltages (u_1, u_2, u_3): to consider them in place of the transformed ones (u_a, u_b), it is necessary to transform them according to the simple law

$$\begin{bmatrix} u_a \\ u_b \end{bmatrix} = T \begin{bmatrix} u_1 \\ u_2 \\ u_3 \end{bmatrix} \tag{6}$$

with

$$T = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & +\frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \tag{7}$$

The control design is based on the definition of three sliding functions which identify the manifold in the system state space such that, when the system trajectory lies on it, the system exhibits the desired dynamics. More precisely, to drive speed and flux tracking errors to zero with exponential law, and to guarantee the symmetry condition to the stator voltages system, the following sliding functions

$$\begin{cases} s_1 & = & k_\omega\tilde{\omega} + \dot{\tilde{\omega}} \\ s_2 & = & k_\Psi\tilde{\Psi} + \dot{\tilde{\Psi}} \\ s_3 & = & \int_0^t \left(u_1 + u_2 + u_3\right) d\tau \end{cases} \tag{8}$$

are selected. To determine the control law that is expected to steer the sliding functions (8) to zero in finite time, one has to consider the dynamics of $s = \left(s_1, s_2, s_3\right)^T$, described by

$$\frac{ds}{dt} = F + Du \tag{9}$$

where $u^T = (u_1, u_2, u_3)$, while vector $F^T = (f_1, f_2, 0)$ and matrix D can be explicitly found by differentiating s_1 and s_2. The components of vector F may be regarded as bounded disturbances, which are in turn continuous functions of motor parameters, speed, rotor fluxes, reference signals and of their first and second time derivatives. Matrix D can be written as

$$D = \begin{bmatrix} k_1 & 0 & 0 \\ 0 & k_2 & 0 \\ 0 & 0 & k_3 \end{bmatrix} \begin{bmatrix} D_1 \\ d \end{bmatrix} \tag{10}$$

with D_1 defined as

$$D_1 \quad = \quad \begin{bmatrix} \frac{1}{J}\frac{M}{L_r}\frac{1}{\sigma L_s} & 0 \\ 0 & \frac{2\alpha M}{\sigma L_s} \end{bmatrix} \begin{bmatrix} -\psi_b & \psi_a \\ \psi_a & \psi_b \end{bmatrix} T \tag{11}$$

and $d = \begin{bmatrix} 1 & 1 & 1 \end{bmatrix}$; k_1, k_2, and k_3 are positive constant design parameters, introduced to virtually increase the control amplitude.

Defining the transformed sliding functions $s^* = \Omega s$, where matrix $\Omega = D^{-1}$ exists everywhere in the system state space, except for $\|\Psi\| = 0$, where $\det D \neq 0$, their time derivative, describing the state motion on the subspace $s^* = 0$, results in

$$\frac{ds^*}{dt} = \Omega F + \frac{d\Omega}{dt} D s^* + u \quad (12)$$

Then, the following switching control law

$$u = -u_0 sign\left(s^*\right) \quad (13)$$

where u_0 is a sufficient high control amplitude, can be chosen to ensure the finite time reaching of $s^* = 0$, [3].

B. The proposed State Feedback Sliding Mode Control

To design a sliding mode control algorithm by assuming the stator currents time derivatives as control inputs, it is first necessary to derive the sliding functions to impose the desired behaviour of speed and flux errors. To this end, let

$$\begin{cases} s_1 &= k_\omega \tilde{\omega} + \dot{\tilde{\omega}} \\ s_2 &= k_\Psi \tilde{\Psi} + \dot{\tilde{\Psi}} \end{cases} \quad (14)$$

with the dynamics of $s^T = (s_1, s_2)$ described by

$$\frac{ds}{dt} = F + D\dot{i} \quad (15)$$

where $\dot{i}^T = \left(\dot{i}_a, \dot{i}_b\right)$ is the two dimensional control. Vector $F^T = (f_1, f_2)$ can be found in the same way as indicated in the previous section, while

$$D = \begin{bmatrix} k_1 & 0 \\ 0 & k_2 \end{bmatrix} \begin{bmatrix} \frac{1}{J}\frac{M}{L_r} & 0 \\ 0 & 2\alpha M \end{bmatrix} \begin{bmatrix} -\psi_b & \psi_a \\ \psi_a & \psi_b \end{bmatrix} \quad (16)$$

with k_1 and k_2 positive constants. By transforming the sliding functions through the use of the matrix $\Omega = D^{-1}$, and remembering that, in this case too, matrix Ω exists when $\|\Psi\| \neq 0$, the state motion on the subspace $s^* = 0$ turns out to be characterized by the equation

$$\frac{ds^*}{dt} = \Omega F + \frac{d\Omega}{dt} D s^* + \dot{i} \quad (17)$$

By choosing the switching control law

$$\dot{i} = -\dot{i}_0 sign\left(s^*\right) \quad (18)$$

for sufficiently high values of the design parameter $\dot{i}_0$ the objective of reaching the manifold $s^* = 0$ in finite time is attained.

In contrast to the nonlinear output feedback control scheme presented in [2], the proposed sliding mode control algorithm does not require the knowledge of the load torque (which is instead assumed to be known in this paper), but only the knowledge of an upperbound of it.

C. The Sliding Mode Adaptive Speed and Flux Observer

An original adaptive speed and flux observer is proposed. Unlike [9] and [10] it does not rely on the equivalent control method [3] according to which unknown quantities are obtained by filtering a discontinuous signal, but on a double sliding mode current estimation.

First, let us suppose that a flux observer is available

$$\begin{cases} \frac{d\hat{\psi}_a}{dt} &= f_{\psi_a} \\ \frac{d\hat{\psi}_b}{dt} &= f_{\psi_b} \end{cases} \quad (19)$$

where f_{ψ_a} and f_{ψ_b} are functions that will be defined later. Now, let us design the first sliding mode current observer

$$\begin{cases} \frac{d\hat{i}_{a_1}}{dt} &= -\beta f_{\psi_a} + \frac{1}{\sigma L_s}\left(u_a - R_s i_a\right) - K_1 sign\left(\tilde{i}_{a_1}\right) \\ \frac{d\hat{i}_{b_1}}{dt} &= -\beta f_{\psi_b} + \frac{1}{\sigma L_s}\left(u_a - R_s i_a\right) - K_1 sign\left(\tilde{i}_{b_1}\right) \end{cases} \quad (20)$$

where, according to [3], vanishing of the estimate errors $\tilde{i}_{a_1} = \hat{i}_{a_1} - i_a$ and $\tilde{i}_{b_1} = \hat{i}_{b_1} - i_b$ is ensured by sufficiently high gain K_1 of the discontinuous signal, introduced to enforce a sliding mode behaviour. By considering the estimate errors dynamics

$$\begin{cases} \frac{d\tilde{i}_{a_1}}{dt} &= -\beta \frac{d\tilde{\psi}_a}{dt} - K_1 sign\left(\tilde{i}_{a_1}\right) \\ \frac{d\tilde{i}_{b_1}}{dt} &= -\beta \frac{d\tilde{\psi}_b}{dt} - K_1 sign\left(\tilde{i}_{b_1}\right) \end{cases} \quad (21)$$

analogously to [2], auxiliary quantities are introduced

$$\begin{cases} z_a &= \tilde{i}_{a_1} + \beta\tilde{\psi}_a \\ z_b &= \tilde{i}_{b_1} + \beta\tilde{\psi}_b \end{cases} \quad (22)$$

which exhibit the dynamics

$$\begin{cases} \frac{dz_a}{dt} &= -K_1 sign\left(\tilde{i}_{a_1}\right) \\ \frac{dz_b}{dt} &= -K_1 sign\left(\tilde{i}_{b_1}\right) \end{cases} \quad (23)$$

and reconstruction of the fluxes estimate errors $\tilde{\psi}_a = \hat{\psi}_a - \psi_a$ and $\tilde{\psi}_b = \hat{\psi}_b - \psi_b$ related to (19) turns out to be feasible, i.e.

$$\begin{cases} \tilde{\psi}_a &= \frac{1}{\beta}\left(z_a - \tilde{i}_{a_1}\right) \\ \tilde{\psi}_b &= \frac{1}{\beta}\left(z_b - \tilde{i}_{b_1}\right) \end{cases} \quad (24)$$

Then, a second sliding mode stator current observer is designed

$$\begin{cases} \frac{d\hat{i}_{a_2}}{dt} &= \frac{1}{\sigma L_s}\left(u_a - R_s i_a\right) - K_2 sign\left(\tilde{i}_{a_2}\right) \\ \frac{d\hat{i}_{b_2}}{dt} &= \frac{1}{\sigma L_s}\left(u_a - R_s i_a\right) - K_2 sign\left(\tilde{i}_{b_2}\right) \end{cases} \quad (25)$$

The estimate errors $\tilde{i}_{a_2} = \hat{i}_{a_2} - i_a$ and $\tilde{i}_{b_2} = \hat{i}_{b_2} - i_b$ dynamics is

$$\left\{\begin{array}{lcl}\frac{d\tilde{\imath}_{a_2}}{dt} & = & +\beta\frac{d\psi_a}{dt}-K_2 sign\left(\tilde{\imath}_{a_2}\right)\\ \frac{d\tilde{\imath}_{b_2}}{dt} & = & +\beta\frac{d\psi_b}{dt}-K_2 sign\left(\tilde{\imath}_{b_2}\right)\end{array}\right. \quad (26)$$

Analogously to the previous case, auxiliary functions are introduced

$$\left\{\begin{array}{lcl}v_a & = & -\tilde{\imath}_{a_2}+\beta\psi_a\\ v_b & = & -\tilde{\imath}_{b_2}+\beta\psi_b\end{array}\right. \quad (27)$$

characterized by the following dynamics

$$\left\{\begin{array}{lcl}\frac{dv_a}{dt} & = & +K_2 sign\left(\tilde{\imath}_{a_2}\right)\\ \frac{dv_b}{dt} & = & +K_2 sign\left(\tilde{\imath}_{b_2}\right)\end{array}\right. \quad (28)$$

So, the second current observer, namely (25), allows as to compute the true rotor flux values

$$\left\{\begin{array}{lcl}\psi_a & = & \frac{1}{\beta}\left(v_a+\tilde{\imath}_{a_2}\right)\\ \psi_b & = & \frac{1}{\beta}\left(v_b+\tilde{\imath}_{b_2}\right)\end{array}\right. \quad (29)$$

Now, relying on knowledge of variables $\tilde{\psi}_a$, $\tilde{\psi}_b$, ψ_a and ψ_b, it is possible to define functions f_{ψ_a} and f_{ψ_b} so that the flux observer (19) can be rewritten as

$$\left\{\begin{array}{lclcl}\frac{d\hat{\psi}_a}{dt} & = & f_{\psi_a} & = & -\hat{\alpha}\psi_a-\hat{\omega}\psi_b+M\hat{\alpha}i_a-K_\psi\tilde{\psi}_a\\ \frac{d\hat{\psi}_b}{dt} & = & f_{\psi_b} & = & -\hat{\alpha}\psi_b+\hat{\omega}\psi_a+M\hat{\alpha}i_b-K_\psi\tilde{\psi}_b\end{array}\right. \quad (30)$$

with $K_\psi > 0$, where $\hat{\omega}$ and $\hat{\alpha}$ are estimated by

$$\left\{\begin{array}{lcl}\frac{d\hat{\omega}}{dt} & = & \frac{1}{J}\frac{M}{L_r}\left(\psi_a i_b-\psi_b i_a\right)-\frac{\Gamma_l}{J}+f_\omega\\ \frac{d\hat{\alpha}}{dt} & = & f_\alpha\end{array}\right. \quad (31)$$

with f_ω and f_α additional terms, to be defined, introduced to impose the desired behaviour to the estimation errors $\tilde{\omega}=\hat{\omega}-\omega$ and $\tilde{\alpha}=\hat{\alpha}-\alpha$, that is

$$\left\{\begin{array}{lcl}\frac{d\tilde{\omega}}{dt} & = & f_\omega\\ \frac{d\tilde{\alpha}}{dt} & = & f_\alpha\end{array}\right. \quad (32)$$

Note that, as previously mentioned, two key assumptions have been taken ito account: as in [2], α is regarded as an unknown but constant quantity, while the load torque is supposed to be known, as assumed in [4].

Relying on (30), the flux estimate errors dynamics turns out to be

$$\left\{\begin{array}{lcl}\frac{d\tilde{\psi}_a}{dt} & = & -\tilde{\alpha}\psi_a-\tilde{\omega}\psi_b+M\tilde{\alpha}i_a-K_\psi\tilde{\psi}_a\\ \frac{d\tilde{\psi}_b}{dt} & = & -\alpha\psi_b+\omega\psi_a+M\tilde{\alpha}i_b-K_\psi\tilde{\psi}_b\end{array}\right. \quad (33)$$

Now, it is possible to select the following Lyapunov function

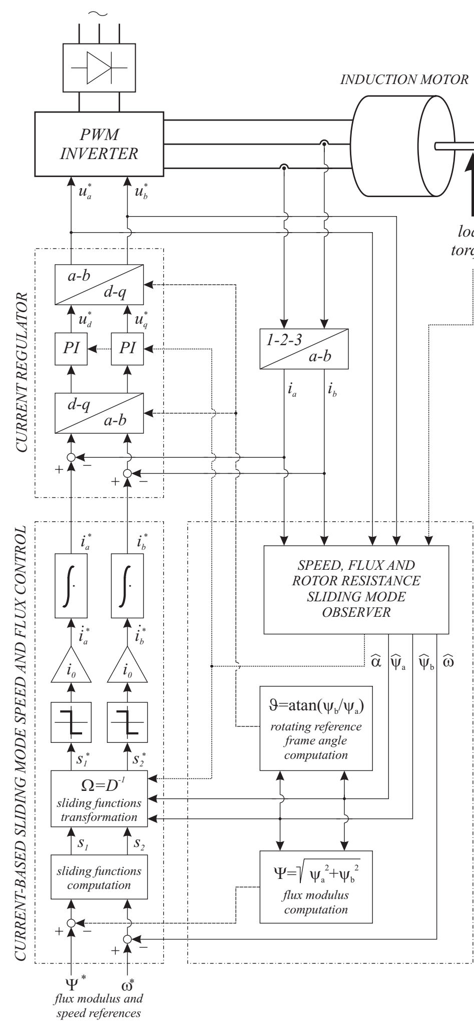

Fig. 1. The proposed induction motor control scheme.

$$V_\omega=\frac{1}{2}\left\{\tilde{\psi}_a^2+\tilde{\psi}_b^2+\frac{1}{\gamma_\omega}\tilde{\omega}^2+\frac{1}{\gamma_\alpha}\tilde{\alpha}^2\right\} \quad (34)$$

in which $\gamma_\omega > 0$ and $\gamma_\alpha > 0$. Its time derivative $\dot{V}_\omega$ is given by

$$\begin{array}{lcl}\dot{V}_\omega & = & \tilde{\psi}_a\dot{\tilde{\psi}}_a+\tilde{\psi}_b\dot{\tilde{\psi}}_b+\frac{1}{\gamma_\omega}\tilde{\omega}\dot{\tilde{\omega}}+\frac{1}{\gamma_\alpha}\tilde{\alpha}\dot{\tilde{\alpha}}\\ & = & -K_\psi\left(\tilde{\psi}_a^2+\tilde{\psi}_b^2\right)\\ & & +\tilde{\omega}\left(\frac{1}{\gamma_\omega}f_\omega-\tilde{\psi}_a\psi_b+\tilde{\psi}_b\psi_a\right)\\ & & +\tilde{\alpha}\left(\frac{1}{\gamma_\alpha}f_\alpha+\tilde{\psi}_a\left(Mi_a-\psi_a\right)+\tilde{\psi}_b\left(Mi_b-\psi_b\right)\right)\end{array} \quad (35)$$

To have $\dot{V}_\omega \leq 0$, one can select

$$\left\{\begin{array}{lcl}f_\omega & = & \gamma_\omega\left(\tilde{\psi}_a\psi_b-\tilde{\psi}_b\psi_a\right)\\ f_\alpha & = & \gamma_\alpha\left(\tilde{\psi}_a\left(\psi_a-Mi_a\right)+\tilde{\psi}_b\left(\psi_b-Mi_b\right)\right)\end{array}\right. \quad (36)$$

Equations (31), with additional terms f_ω and f_α defined in (36), provide an estimation law for the mechanical speed ω and an adaptive law for the unknown parameter α. Convergence of the flux observer (30) is so guaranteed.

The Fig. 1 shows the proposed induction motor control scheme with full details.

D. Some comments on convergence

To guarantee the convergent behaviour of the update law of α, it is necessary to show that the signal which drives such a law verifies the persistency of excitation condition [12]. To this aim, the second equation in (31) can be put in the form

$$\begin{aligned} \frac{d\hat{\alpha}}{dt} &= \gamma_\alpha \begin{bmatrix} (\psi_a - M i_a) & (\psi_b - M i_b) \end{bmatrix} \begin{bmatrix} \tilde{\psi}_a \\ \tilde{\psi}_b \end{bmatrix} \\ &= \gamma_\alpha \Gamma(t) \begin{bmatrix} \tilde{\psi}_a \\ \tilde{\psi}_b \end{bmatrix} \end{aligned} \tag{37}$$

Persistency of excitation is then guaranteed provided that

$$\int_t^{t+T} \Gamma(\tau) \Gamma^T(\tau)\, d\tau \tag{38}$$

is positive definite for $T > 0$ and for any $t \geq 0$. This condition is satisfied, since it results $\forall t \geq 0$

$$\int_t^{t+T} \left((\psi_a - M i_a)^2 + (\psi_b - M i_b)^2 \right) d\tau \geq 0 \tag{39}$$

A similar analysis may be performed about ω.

IV. Simulation Examples

A. Simulation setup

To validate the proposed control algorithm (18) and the speed and flux adaptive observer, simulations have been carried out by use of Matlab and Simulink, adopting the same parameters of the experimental setup shown in [2], in which a 600 W one pole pair induction motor with a rated speed of 1000 rpm is used.

The main purposes were to inspect both performances and robustness properties in reference tracking and observation accuracy. About the control algorithm, another task is to verify that the limit imposed to the maximum value of the stator current time derivatives does not compromise the dynamical performances during transients.

The speed and flux modulus references and the load torque profile are shown in Fig. 2: both the first and the second time derivatives of speed and flux reference signals are bounded. The simulation here shown has been carried out with a value of the rotor resistance equal to 150% of the nominal one.

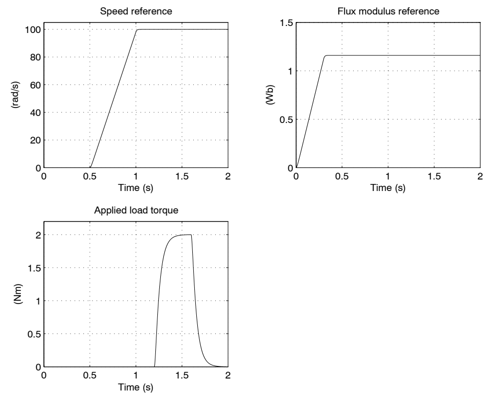

Fig. 2. Speed and rotor flux reference signals and load torque profile in simulations.

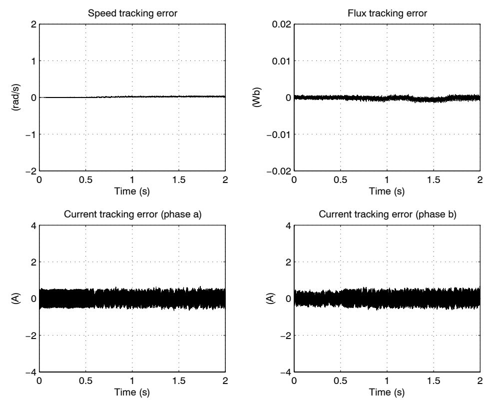

Fig. 3. Speed and rotor flux modulus tracking errors; tracking error of the stator current.

B. Simulation results

The control algorithm performances are illustrated in Fig. 3: current tracking error due to the current regulator is also shown. Fig. 4 reports, in the restricted time interval [0.95, 1.05] s, both the discontinuous waveform (phase a) of the control signal, and the stator current i_a of the motor, thus letting us appreciate the filtering action of the integrators, of the current loop and of the same motor. Moreover, the limit imposed on the time derivative of the stator current (2000 A/s) seems not to affect negatively the speed regulation during load torque transients.

The adaptive sliding mode speed and flux observer proves to be fast and accurate: the previous analysis demonstrates that speed and flux regulation seems not to be affected by the presence of the observer in the control scheme. Fig. 5 shows that speed and flux modulus estimation performances are satisfactory, even before that the convergence of the parameter α estimation is verified, proving good robustness of the speed and flux observer. Both current and flux estimate errors are quickly steered to zero, as shown in Fig. 6 (only performances of the first sliding mode current observer are here reported).

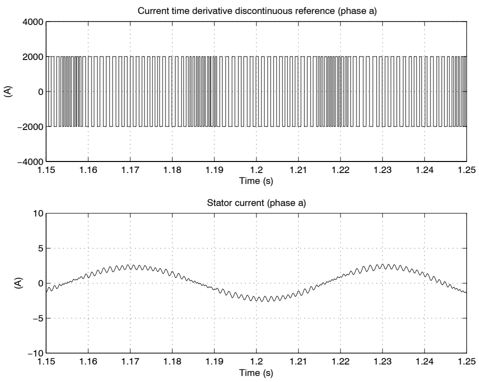

Fig. 4. The switching control input, i.e. the time derivative of the stator current (phase a) and the measured current i_a, showing the high harmonics filtering.

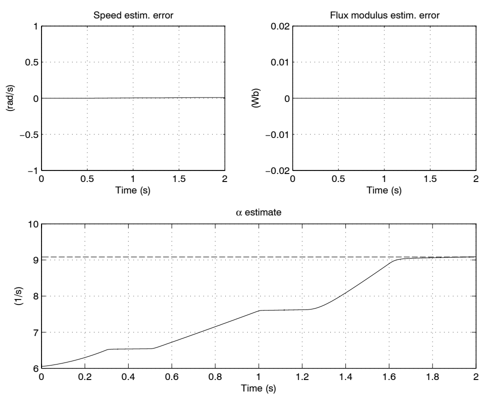

Fig. 5. Speed and flux modulus tracking performances, and α estimation, with the new Adaptive Sensorless Sliding Mode Observer.

V. CONCLUSIONS

In this paper a new Adaptive Sensorless Sliding Mode Control algorithm for the induction motor is proposed. The control strategy sets a limit to the maximum value of the time derivatives of the stator currents, assumed as discontinuous control inputs, in order to prevent excessive mechanical stress of the machine. The novel Adaptive Sliding Mode Speed and Flux Observer, based on a double robust stator current estimation, provides a fast and precise adaptation of both the mechanical speed and the rotor resistance.

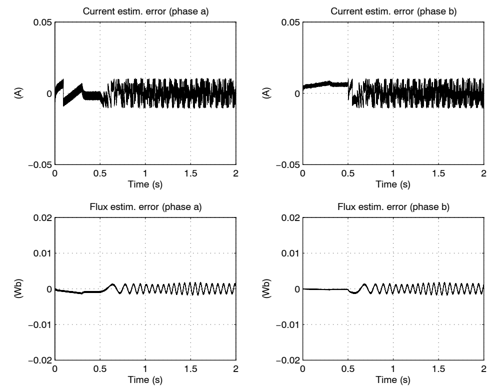

Fig. 6. Stator currents and rotor fluxes components estimate errors.

REFERENCES

[1] W. Leonhard, *Control of Electrical Drives*, 2nd ed. Berlin, Germany: Springer-Verlag, 1996.

[2] R. Marino, S. Peresada, and P. Tomei, "Output feedback control of current-fed induction motors with unknown rotor resistance," *IEEE Trans. Contr. Systems Technology*, vol. 4, no. 4, pp. 336–347, July 1996.

[3] V. I. Utkin, *Sliding Modes in Control and Optimization*. Berlin: Springer-Verlag, 1992.

[4] R. Marino, P. Tomei, and C. M. Verrelli, "A new global control scheme for sensorless current-fed induction motors," in *15th IFAC World Congress*, Barcelona, Spain, 2002.

[5] R. Marino, S. Peresada, and P. Tomei, "Global adaptive output feedback control of induction motors with uncertain rotor resistance," *IEEE Trans. Automat. Contr.*, vol. 44, no. 5, pp. 967–983, May 1999.

[6] C. Aurora, E. Bassi, and A. Ferrara, "Sliding mode control of current-fed induction motors with unknown rotor resistance," in *Proc. Automation and Decision-Making International Conference BIAS 2000*, Milan, Italy, 2000.

[7] C. Aurora, A. Ferrara, and A. Levant, "Speed regulation of induction motors: A sliding mode observer-differentiator based control scheme," in *Proc. 40th IEEE Conference on Decision and Control*, Orlando, Florida, 2001.

[8] R. Marino, S. Peresada, and P. Tomei, "On-line stator and rotor resistance estimation for induction motors," *IEEE Trans. Contr. Systems Technology*, vol. 8, no. 3, pp. 570–579, May 2000.

[9] Z. Yan, C. Jin, and V. I. Utkin, "Sensorless sliding-mode control of induction motors," *IEEE Trans. Ind. Electron.*, vol. 47, pp. 1286–1297, 2000.

[10] A. Derdiyok, M. K. Gven, H. Rehman, N. Inanc, and L. Xu, "Design and implementation of a new sliding-mode observer for speed-sensorless control of induction machine," *IEEE Trans. Ind. Electron.*, vol. 49, no. 5, pp. 1177–1182, 2002.

[11] G. Bartolini, A. Damiano, G. Gatto, I. Marongiu, A. Pisano, and E. Usai, "Robust speed and torque estimation in electrical drives by second-order sliding modes," *IEEE Trans. Contr. Syst. Technol.*, vol. 11, no. 1, pp. 217–220, 2003.

[12] K. S. Narendra and A. M. Annaswamy, *Stable Adaptive Systems*. Englewood Cliffs, NJ: Prentice Hall, 1989.