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Adaptative Variable Structure Control for an Online Tuning Direct Vector Controlled Induction Motor Drives

Sbita Lassâad, Zaltni Dalila and Abdelkrim Mohamed Naceur
Research Unit of Modeling, Analysis and Control Systems,
Department of Electrical and Automatic Engineering, National Engineering School,
Gabes 6029, Tunisia

Abstract: This study demonstrates that high performance speed control can be obtained by using an adaptative sliding mode control method for a direct vector controlled Squirrel Cage Induction Motor (SCIM). In this study a new method of designing a simple and effective adaptative sliding mode rotational speed control law is developed. The design includes an accurate sliding mode flux observation from the measured stator terminals and rotor speed. The performance of the Direct Field-Orientation Control (DFOC) is ensured by online tuning based on a Model Reference Adaptative System (MRAS) rotor time constant estimator. The control strategy is derived in the sense of Lyapunov stability theory so that the stable tracking performance can be guaranteed under the occurrence of system uncertainties and external disturbances. The proposed scheme is a solution for a robust and high performance induction motor servo drives. Simulation results are provided to validate the effectiveness and robustness of the developed methodology.

Key words: Adaptative, sliding mode, observer, induction motor, field oriented control, MRAS estimator, rotor time constant

INTRODUCTION

The SCIM are widely used in industry due to their relatively low cost and high reliability. However, induction motors constitute a theoretically challenging control problem, since the dynamical system is highly coupled, nonlinear and multivariable. The field orientation control is the most commonly used technique for high performance SCIM drive systems. The rotor speed is asymptotically decoupled from the rotor flux and the speed is linearly related to the torque current. Thus, The SCIM processes the same behavior of a separately excited DC motor (Utkin and Sbanovic, 1999). In general, the DFOC performance is very sensitive to parameters variation, especially the rotor time constant (Constantines, 2006; Sbita and Ben Hamed, 2007a). An online adaptation of the rotor time constant is necessary to keep the machine field oriented. In order to ensure robustness under the occurrence of uncertainties, many robust control and observation approaches have been designed (Barambones and Garrido, 2004; Lai and Shyn, 2005; Wai, 2000).

During the last few years, significant interest in Sliding Mode Control (SMC) and observation of SCIM have been generated by researchers and application engineers as well (Benchaib *et al*., 1999; Cheng *et al*., 2007; Zheng *et al*., 2006; Gallah *et al*., 2007). The most salient feature of the SMC is the existence of the motion in selected manifolds in state space. Such motion results in a system performance that includes disturbance rejection and insensitivity to parameters variations (Utkin and Sbanovic, 1999; Young *et al*., 1999). The performance of the SMC is guaranteed by respecting Lyaounov stability conditions (Zaltni *et al*., 2006). In order to meet those conditions a suitable choice of the sliding gain should be made to compensate for uncertainties. In order to select the sliding gain, an upper bound of the parameter variations, external disturbance, unmodelled dynamics and noise magnitude should be known (Lai and Shyn, 2005; Zheng *et al*., 2006; Ben Hamed and Sbita, 2007). However, in practical applications those bounds are generally unknown or at least difficult to calculate. To overcome this problem, an effective way is to choose a larger sliding gain, but it will result in excessive chattering control efforts and may exite high frequency unstable dynamics (Zaltni, 2007; Gallah *et al*., 2007). The concept of an adaptative sliding gain can be introduced to avoid these drawbacks and to improve stability.

Corresponding Author: Sbita Lassâad, Research Unit of Modeling, Analysis and Control Systems, Department of Electrical and Automatic Engineering, National Engineering School, Gabes 6029, Tunisia

Induction motor model: By defining $[v_{ds}\ v_{qs}]^T$ as the vector of stator voltages, $[i_{ds}\ i_{qs}]^T$ as the vector of stator currents, $[\phi_{dr}\ \phi_{qr}]^T$ as the vector of rotor fluxes, ω_s as the synchronous angular speed and ω as the motor electrical angular speed, the dynamic model of the induction motor in a synchronous rotating reference frame (d_q) is given by the following equation (Sbita and Ben Hamed, 2007b).

$$\begin{cases} \frac{di_{ds}}{dt} = a_1 i_{ds} + a_2\phi_{qr}\omega + a_3\phi_{dr} + \omega_s i_{qs} + \eta v_{ds} \\ \frac{di_{qs}}{dt} = a_1 i_{qs} - a_2\phi_{dr}\omega + a_3\phi_{qr} - \omega_s i_{ds} + \eta v_{qs} \\ \frac{d\phi_{dr}}{dt} = -a_4\phi_{dr} - \phi_{qr}(\omega - \omega_s) + a_5 i_{ds} \\ \frac{d\phi_{qr}}{dt} = -a_4\phi_{qr} + \phi_{dr}(\omega - \omega_s) + a_5 i_{qs} \\ \frac{d\omega}{dt} = \frac{P}{J}(T_e - T_l) - \frac{f}{J}\omega \end{cases} \quad (1)$$

Where:

$$a_1 = -\frac{1}{\sigma L_s}(R_s + \frac{R_r L_m{}^2}{L^2{}_r})$$

$$a_2 = \frac{L_m}{\sigma L_s L_r}$$

$$a_3 = \frac{L_m R_r}{\sigma L_s L^2{}_r}$$

$$a_4 = \frac{R_r}{L_r}$$

$$a_5 = \frac{R_r L_m}{L_r}$$

$$\eta = \frac{1}{\sigma L_s}$$

T_e s is the electromagnetic torque, T_l is the load torque, P is the number of pole pairs, f is the friction coefficient, J is the moment inertia;

$$\sigma = 1 - \frac{L_m{}^2}{L_s L_r}$$

is the motor leakage coefficient, L_m, L_r and L_s are mutual, rotor and stator, inductances, R_r and R_s are rotor and stator resistance, respectively and $T_r = \frac{L_r}{R_r}$ is the rotor time constant.

The subscripts s and r denote parameters and variables respectively referred to the stator and rotor and d and q denote the (d_q) axis components in the rotating reference frame.

Sliding mode control law synthesis: The electromagnetic torque can be expressed in the following form:

$$T_e = \frac{PL_m}{L_r}(i_{qs}\phi_{dr} - i_{ds}\phi_{qr}) \quad (2)$$

This shows the coupling between the torque and flux commands. A DFOC is used to solve this problem. The application of DFOC strategy consists in the orientation of the flux vector along the d axis, which leads:

$$\phi^{(d,q)} = (\phi_r, 0),\ \phi_r \text{ the nominal rotor flux} \quad (3)$$

Under field-oriented assumptions, the electromagnetic torque (2) and the rotor flux dynamic expressed in (1) can be simplified as:

$$T_e = K_T i_{sq}{}^* \text{ with } K_T = \frac{PL_m}{Lr}\phi^*_{rd} \quad (4)$$

$$\phi_r{}^* = \frac{L_m}{1+\tau_r s} i_{ds}{}^* \quad (5)$$

Where, K_T is the torque constant, ϕ^*_{rd} is the flux reference, i_{ds}* and i_{qs}* denotes the flux and torque current commands, s denoting the Laplace operator.

In order to guarantee that the synchronous angular speed ω_s of the (d_q) axis reference frame is effectively the same as the rotating reference frame, the following relation should always be verified (Ben Hamed and Sbita, 2006).

$$\omega_s = \omega + \frac{L_m i_{sq}}{\tau_r \phi_{rd}} \quad (6)$$

The flux current command i_{ds}* is calculated using the following equation.

$$i^*_{sd} = K_{i\phi}\int_0^t (\phi_{rd} - \phi^*_{rd})(\tau)d\tau + K_{p\phi}(\phi_{rd} - \phi^*_{rd}) + \frac{\tau_r}{M}\dot{\phi}^*_{rd} + \frac{1}{M}\phi^*_{rd}; \quad (7)$$

Since the flux is constant in the machine, we can now calculate the speed current command i_{qs}* using the sliding mode technique (Barambones *et al*., 2004; Yue *et al*., 2005).

Substituting Eq. 5 into 2, the rotor speed dynamic of the IM drive system is:

$$\frac{d\omega}{dt} = hi_{sq}{}^* - a\omega - bT_l \quad (8)$$

Where:

$$h = \frac{PK_t}{J}$$

$$a = \frac{f}{J}$$

$$b = \frac{P}{J}$$

Now we are going to consider the previous Eq. 8 with uncertainties as follows:

$$\frac{d\omega}{dt} = (h + \Delta h)i_{sq}^* - (a + \Delta a)\omega - (b + \Delta b)T_l \quad (9)$$

Where, Δh, Δa, Δb denote the uncertainties of the terms h, a and b, respectively. The tracking speed error is defined as:

$$e(t) = \omega - \omega^* \quad (10)$$

The control law should ensure that the tracking speed error e (t) tends to zero as the time tends to infinity.

The dynamic of the rotor speed error is given by:

$$\dot{e}(t) = \dot{\omega}(t) - \dot{\omega}^*(t) \quad (11)$$

Tacking into account the Eq. 9, the previous equation becomes:

$$\dot{e}(t) = -ae(t) + u(t) + \varepsilon(t) \quad (12)$$

Where:

$$u(t) = hi_{sq} - a\omega^* - \dot{\omega}^* \quad (13)$$

and $\varepsilon(t) = -\Delta a\omega - (b+\Delta b)T_l + \Delta hi_{sq}$ is the term containing uncertainties.

Assumption: If $\varepsilon(t) = 0$ and if we take $u(t) = ke(t)$, where, k is a constant that will be defined later, then the Eq. 12 becomes:

$$\dot{e}(t) = (k - a)e(t) \quad (14)$$

The previous equation shows that the tracking speed error converges to zero exponentially as the time tends to infinity when (k-a) is strictly negative. It is the desired behavior of the rotor speed tracking error. This desired behavior could be obtained by choosing the sliding surface as follows:

$$S(t) = e(t) - \int_0^t (k - a)e(\tau)d\tau \quad (15)$$

In fact, when the sliding mode occurs on the sliding surface then, $\dot{S}(t) = 0$ yields to $\dot{e}(t) - (k - a)e(t) = 0$.

The variable structure speed controller is designed as:

$$u(t) = u_{eq} - M\,\mathrm{sgn}(S) \quad (16)$$

Where, $u_{eq} = ke(t)$ is the equivalent control obtained when, $\dot{S}(t) = 0$, $M\,\mathrm{sgn}(S)$ is a high frequency discontinuous term that guarantees the robustness of the control law and maintain the phase trajectories on the sliding surface and M is the adaptative switching gain of the sign function which is chosen so as to guarantee the stability of the closed loop system. Let us define the lyapunov function candidate and its time derivative as:

$$V_1(t) = \frac{1}{2}S^2(t)\text{: Positive definite and}$$

$$\dot{V}_1(t) = \dot{S}(t)S(t) = -[M - |\varepsilon|]|S| \le 0$$

It is clear that $\dot{V}_1(t)$ is negative definite as long as the following condition is satisfied: $M \ge |\varepsilon|$.

For satisfying the previous condition, the load torque value and mechanical parameter uncertainties should be known (Loukianov *et al*., 2001). However, in practical applications, the load torque and the uncertainties are generally unknown or at least difficult to determine precisely. One effective way is to choose a large value of the sliding gain; therefore, it will result in a chattering phenomenon, which is considered to be the most serious obstacle in the application of sliding mode control (Cheng *et al*., 2007; Chen and Peng, 2005). In order to overcome this problem and since the sliding surface is influenced by choice of the sliding gain, then the proposed adaptative sliding mode control is expressed as follows:

$$\dot{M} = \gamma|S| \quad \text{with} \quad M(0) = 0 \quad (17)$$

The gain γ is a positive constant.

Therefore, by choosing k so as (k-a) is strictly negative, all trajectories converges to:

$$S(t) = e(t) - \int_0^t (k - a)e(\tau)d\tau$$

and remains on this surface

$$S(t) = \dot{S}(t) = 0$$

Then the dynamic behavior of the rotor tracking speed error $\dot{e}(t) = -ae(t) + u(t) + \varepsilon(t)$ becomes $\dot{e}(t) = (k-a)e(t)$ which converges exponentially to zero as the time tends to infinity.

Finally, the torque current commands $i_{qs}*$ can be obtained directly by substituting the Eq. 13 in the Eq. 16:

$$i_{sq}* = \frac{1}{h}[ke - M\,\mathrm{sgn}(S) + a\omega* + \dot{\omega}*] \quad (18)$$

In order to force i_{ds} and I_{qs} currents to their references $i_{ds}*$ and $i_{qs}*$, a current loop with fast dynamics (PI controllers) is used. Then, the voltage commands can be written as:

$$\begin{cases} V_{sd}* = K_{ivd}\int_0^t (i*_{sd} - i_{sd})dt + K_{pvd}(i*_{sd} - i_{sd}) \\ V_{sq}* = K_{ivq}\int_0^t (i*_{sq} - i_{sq})dt + K_{pvq}(i*_{sq} - i_{sq}) \end{cases} \quad (19)$$

where:

K_{ivd}, K_{ivq} = The constants of integral action

K_{pvd}, K_{pvq} = The constants of proportional action.

Sliding mode observer: Considering only the first four equations of the induction motor model (1) in the stationary (α, β) coordinate system and assuming that the only output measured variables are the stator currents and the rotor speed. The observer model is a copy of the original system, which has corrector gains with switching terms (Zaltni, 2007; Gallah *et al*., 2007; Ben Hamed and Sbita, 2006):

$$\begin{cases} \frac{d\hat{i}_{s\alpha}}{dt} = -a_1\hat{i}_{s\alpha} + a_2\hat{\phi}_{r\beta}\omega + a_3\hat{\phi}_{r\alpha} + \eta v_{s\alpha} + K_1 \\ \frac{d\hat{i}_{s\beta}}{dt} = -a_1\hat{i}_{s\beta} - a_2\hat{\phi}_{r\alpha}\omega + a_3\hat{\phi}_{r\beta} + \eta v_{s\beta} + K_2 \\ \frac{d\hat{\phi}_{r\alpha}}{dt} = -a_4\hat{\phi}_{r\alpha} - \hat{\phi}_{r\beta}\omega + a_5\hat{i}_{s\alpha} + K_3 \\ \frac{d\hat{\phi}_{r\beta}}{dt} = -a_4\hat{\phi}_{r\beta} + \hat{\phi}_{r\alpha}\omega + a_5\hat{i}_{s\beta} + K_4 \end{cases} \quad (20)$$

Where:

$\hat{i}_{s\alpha}$, $\hat{i}_{s\beta}$, $\hat{\phi}_{s\alpha}$ et $\hat{\phi}_{r\beta}$ = The estimates.

K_1, K_2, K_3 and K_4 = The observer gains.

The estimation error dynamics are given by:

$$\begin{cases} \frac{de_1}{dt} = -a_1e_1 + a_2e_4\omega + a_3e_3 + K_1 \\ \frac{de_2}{dt} = -a_1e_2 - a_2e_3\omega + a_3e_4 + K_2 \\ \frac{de_3}{dt} = -a_4e_3 - e_4\omega + a_5e_1 + K_3 \\ \frac{de_4}{dt} = -a_4e_4 + e_3\omega + a_5e_2 + K_4 \end{cases} \quad (21)$$

Where:

$e_1 = \hat{i}_{s\alpha} - i_{s\alpha}$

$e_2 = \hat{i}_{s\beta} - i_{s\beta}$

$e_3 = \hat{\phi}_{r\alpha} - \phi_{r\alpha}$

$e_4 = \hat{\phi}_{r\beta} - \phi_{r\beta}$

$$\begin{bmatrix} K_1 \\ K_2 \end{bmatrix} = \begin{bmatrix} -\lambda_1 \mathrm{sign}(e_1) \\ -\lambda_2 \mathrm{sign}(e_2) \end{bmatrix} \text{ and } \begin{bmatrix} K_3 \\ K_4 \end{bmatrix} = -\Delta \begin{bmatrix} K_1 \\ K_2 \end{bmatrix} (\Delta \in \mathbb{R}^{2*2}) \quad (22)$$

Let's define the sliding surface of the observer as follows:

$$S_{ob} = [s_{ob1} \quad s_{ob2}]^T = [e_1 \quad e_2]^T$$

The stability analysis of the system (21) consists of determining K_1 and K_2 to ensure the reachability of the sliding surface S_{ob} = 0. Thereafter, K_3 and K_4 are determined so that the reduced-order system obtained when $S_{ob} = \dot{S}_{ob} = 0$ is locally stable. When the sliding mode occurs then, $S_{ob} = \dot{S}_{ob} = 0$ and the equation system (21) becomes:

$$\begin{cases} 0 = a_2e_4\omega + a_3e_3 + K_{1eq} \\ 0 = -a_2e_3\omega + a_3e_4 + K_{2eq} \end{cases} \quad (23)$$

$$\begin{cases} \frac{de_3}{dt} = -a_4e_3 - e_4\omega + K_{3eq} \\ \frac{de_4}{dt} = -a_4e_4 + e_3\omega + K_{4eq} \end{cases} \quad (24)$$

From the Eq. 23, we deduce:

$$\begin{bmatrix} K_{1eq} \\ K_{2eq} \end{bmatrix} = -\begin{bmatrix} a_3 & a_2\omega \\ -a_2\omega & a_3 \end{bmatrix} \begin{bmatrix} e_3 \\ e_4 \end{bmatrix} = -\Gamma \begin{bmatrix} e_3 \\ e_4 \end{bmatrix} \quad (25)$$

From Eq. 24, if we choose K_{3eq} and K_{4eq} as follows:

$$\begin{bmatrix} K_{3eq} \\ K_{4eq} \end{bmatrix} = -\begin{bmatrix} \beta_1 & -\omega \\ \omega & \beta_2 \end{bmatrix}\begin{bmatrix} e_3 \\ e_4 \end{bmatrix} = -\psi\begin{bmatrix} e_3 \\ e_4 \end{bmatrix} \quad (26)$$

The flux error dynamic becomes:

$$\begin{bmatrix} \dot{e}_3 \\ \dot{e}_4 \end{bmatrix} = \begin{bmatrix} -(a_4+\beta_1) & 0 \\ 0 & -(a_4+\beta_2) \end{bmatrix}\begin{bmatrix} e_3 \\ e_4 \end{bmatrix} \quad (27)$$

By choosing β_1 and β_2 positive large sufficiently, e_3 and e_4 converge exponentially to zero as long as the time tends to infinity. Since $\det(\Gamma) = a_3 2+(a_2\omega)^2 \neq 0$, then K_{3eq} and k_{4eq} can be deduced from Eq. 25 and 26:

$$\begin{bmatrix} K_{3eq} \\ K_{4eq} \end{bmatrix} = -\psi\Gamma^{-1}\begin{bmatrix} K_{1eq} \\ K_{2eq} \end{bmatrix} = -\Delta\begin{bmatrix} K_{1eq} \\ K_{2eq} \end{bmatrix}, \text{ with } \Delta = \psi\Gamma^{-1} \text{ then:}$$

$$\begin{bmatrix} K_{3eq} \\ K_{4eq} \end{bmatrix} = -\Delta\begin{bmatrix} K_{1eq} \\ K_{2eq} \end{bmatrix} = -\Delta\Gamma\begin{bmatrix} e_1 \\ e_2 \end{bmatrix} \quad (28)$$

The stability of the overall observer structure is guaranteed through the stability of the current observer The Lyapunov function for the proposed sliding mode observer is chosen as:

$$V_2 = \frac{1}{2}s_{ob1}^2 + \frac{1}{2}s_{ob2}^2$$

The Lyapunov function V_2 is positive definite. This satisfies the first Lyapunov stability condition. The second condition is that the derivative of the Lyapunov function must be less than zero (Cheng *et al*., 2007; Zaltni, 2007).

$$\text{In fact: } \dot{V}_2 = -a_1(s_{ob1}^2 + s_{ob2}^2) - \lambda_1|s_{ob1}| - \lambda_2|s_{ob2}| \leq 0 \,\forall \lambda_1, \lambda_2 \geq 0.$$

MRAS rotor time constant estimator: Here, a Model Reference Adaptative System (MRAS) observer for the inverse of the rotor time constant is designed. This observer uses for flux the Voltage Model (VM) as the reference and the Current Model (CM) as the adjustable model (Schauder, 1989; Sbita and Ben Hamed, 2007b). The VM computes the stator and rotor fluxes in the stationary reference frame as:

$$\begin{cases} \phi_s = \int (Vs - R_s i_s)dt = \int e_s dt \\ \phi_{r\ \alpha\beta}^{ref} = \frac{L_r}{L_m}(\phi_s - \sigma L_s i_s) \end{cases} \quad (29)$$

Where, $\phi_s, \phi_{r\ \alpha\beta}^{ref}$ are the stator and rotor-flux vectors. The equations of the (CM) observer are:

$$\begin{cases} \frac{d\hat{\phi}_{\alpha r}}{dt} = -\frac{1}{T_r}\hat{\phi}_{\alpha r} - \hat{\phi}_{\beta r}\,\omega + \frac{L_m}{T_r}i_{\alpha s} \\ \frac{d\hat{\phi}_{\beta r}}{dt} = -\frac{1}{Tr}\hat{\phi}_{\beta r} + \hat{\phi}_{\alpha r}\,\omega + \frac{L_m}{Tr}i_{\beta s} \end{cases} \quad (30)$$

Now, let us define the error vector as follows:

$$\bar{\varepsilon} = \bar{\phi}_r - \hat{\phi}_r \quad (31)$$

It's time derivative is given by:

$$\dot{\bar{\varepsilon}} = [A]\bar{\varepsilon} - [W] \quad (32)$$

Where:

$$A = \begin{bmatrix} \frac{-1}{T_r} & -\omega \\ \omega & \frac{-1}{T_r} \end{bmatrix}, \quad W = \left(\frac{1}{T_r} - \frac{1}{\hat{T}_r}\right)(\hat{\phi}_r - L_m i_s)$$

The matrix A can be considered as the complex pole of the dynamic error of the linear system. This pole has a negative real part then it's a stable system.

The term W should tend to zero. Using the proposed adaptation law, the estimated value of $\frac{1}{T_r}$ is given by the following algorithm.

$$\frac{1}{\hat{T}_r} = K_p[\varepsilon_\alpha(L_m i_{\alpha s} - \hat{\phi}_{r\alpha}) + \varepsilon_\beta(L_m i_{s\beta} - \hat{\phi}_{r\beta})] + K_i\int_0^t [\varepsilon_\alpha(L_m i_{\alpha s} - \hat{\phi}_{r\alpha}) + \varepsilon_\beta(L_m i_{s\beta} - \hat{\phi}_{\beta q})]d\tau \quad (33)$$

The bloc scheme of the proposed estimator algorithm is shown in Fig. 1.

Simulation results: The performance of the proposed control law was investigated by simulation using Matlab/Simulink package for the dynamic model of a SCIM written in the rotating reference frame. The used SCIM characteristics: Rated power, voltage, current, number of pole pairs, frequency and speed have, respectively the following values: 1.5 Kw, (220-380)v, (6.4-3.7)A, 2, 50 Hz, 1410 r. mn^{-1}. The motor parameters:

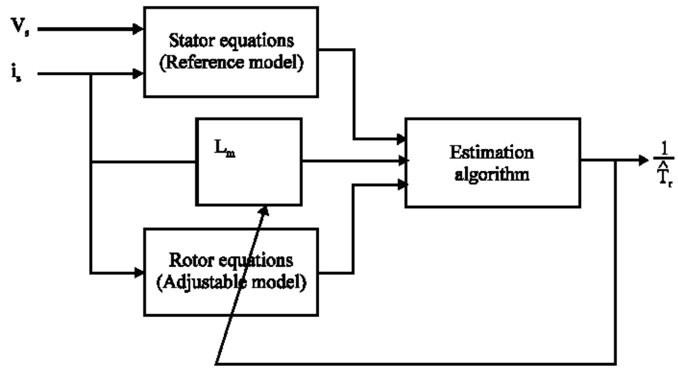

Fig. 1: MRAS rotor time constant estimator scheme

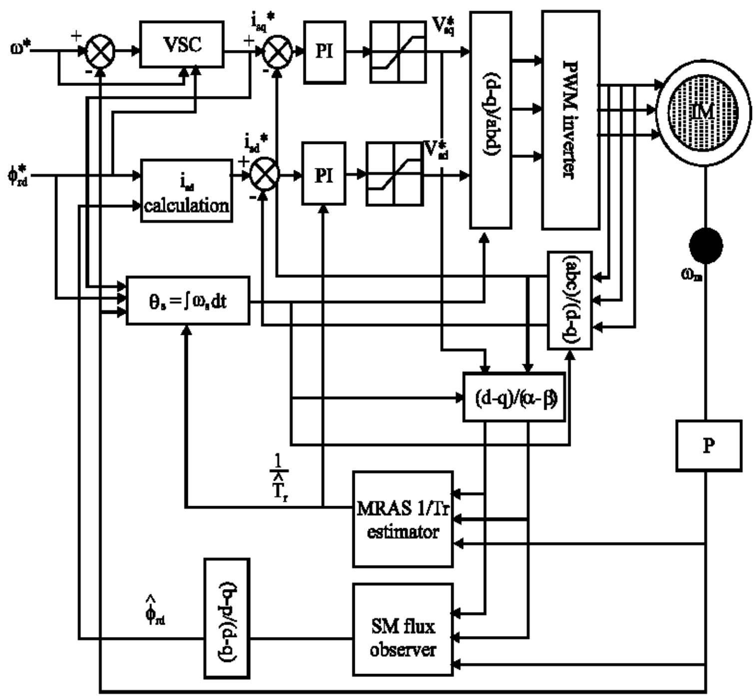

Fig. 2: Control scheme of the adaptative VSC of SCIM

R_s, R_r, L_s, L_r, L_m, J and f have, respectively the following nominal values: 2.3Ω, 1.55Ω, 0.261H, 0.261H, 0.249H, 0.0076 kg m^2, 0.007 Nm sec rad^{-1}. The controllers parameter values are: k = -400, γ = 30, K_{pvd} = 10, K_{ivd} = 1000, K_{pvq} = 0.5, K_{ivq} = 1000, $K_{i\phi}$ = 5, $K_{i\phi}$ = 10 and the observer parameter values are fixed as: β_1 = 30, β_2 = 30, λ_1 = 10 and λ_2 = 10.

The MRAS rotor time constant estimator is presented in Fig. 1 and the complete control scheme is summarized in Fig. 2. The bloc VSC represents the proposed adaptative variable structure controller and it is implemented by Eq. 15, 17 and 18. The bloc of θ_s calculation represents the flux angle which is implemented by the integration of the Eq. 6. The bloc i_{ds} calculation gives the reference current i_{ds}* using Eq. 7. The bloc MRAS_1/Tr estimator represents the MRAS rotor time estimator and it concerns the integration of Eq. 29, 30 and 33. The bloc SM flux observer deals with the sliding mode rotor flux observer which Eq. 20, 22 and 28 are implemented. An extensive simulation work was performed and selected results are presented here to illustrate the most important finding in this study.

Figure 3 shows the motor dynamic responses with no external disturbances and with no uncertainties. Figure 4 gives the motor dynamic responses with the application of a load torque of 2 N. m on time t = 2 sec, 4 N. m on time t = 4 sec and 6 N. m on time t = 8 sec and a variation of R_r up to 50% of its nominal value on time t = 3 sec.

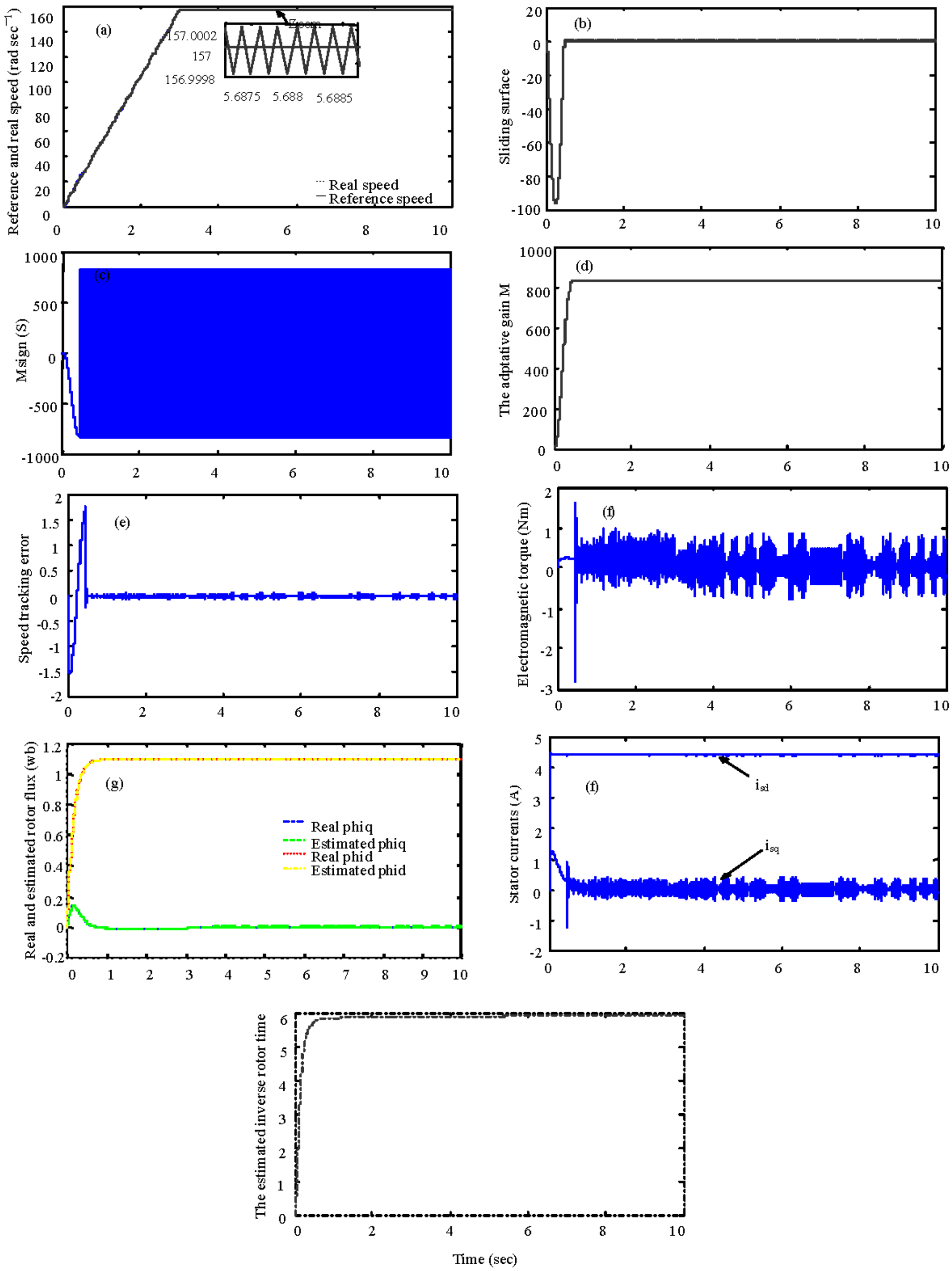

Fig. 3: (a) Real and reference speed (rad/s) (b) Sliding surface (c) Msign(S) (d) The adaptative gain (e) Speed tracking error (f) Electromagnetic torque (Nm) (g) Real and estimated rotor flux (wb) (h) Stator current (A) (i) The estimated inverse rotor time constant

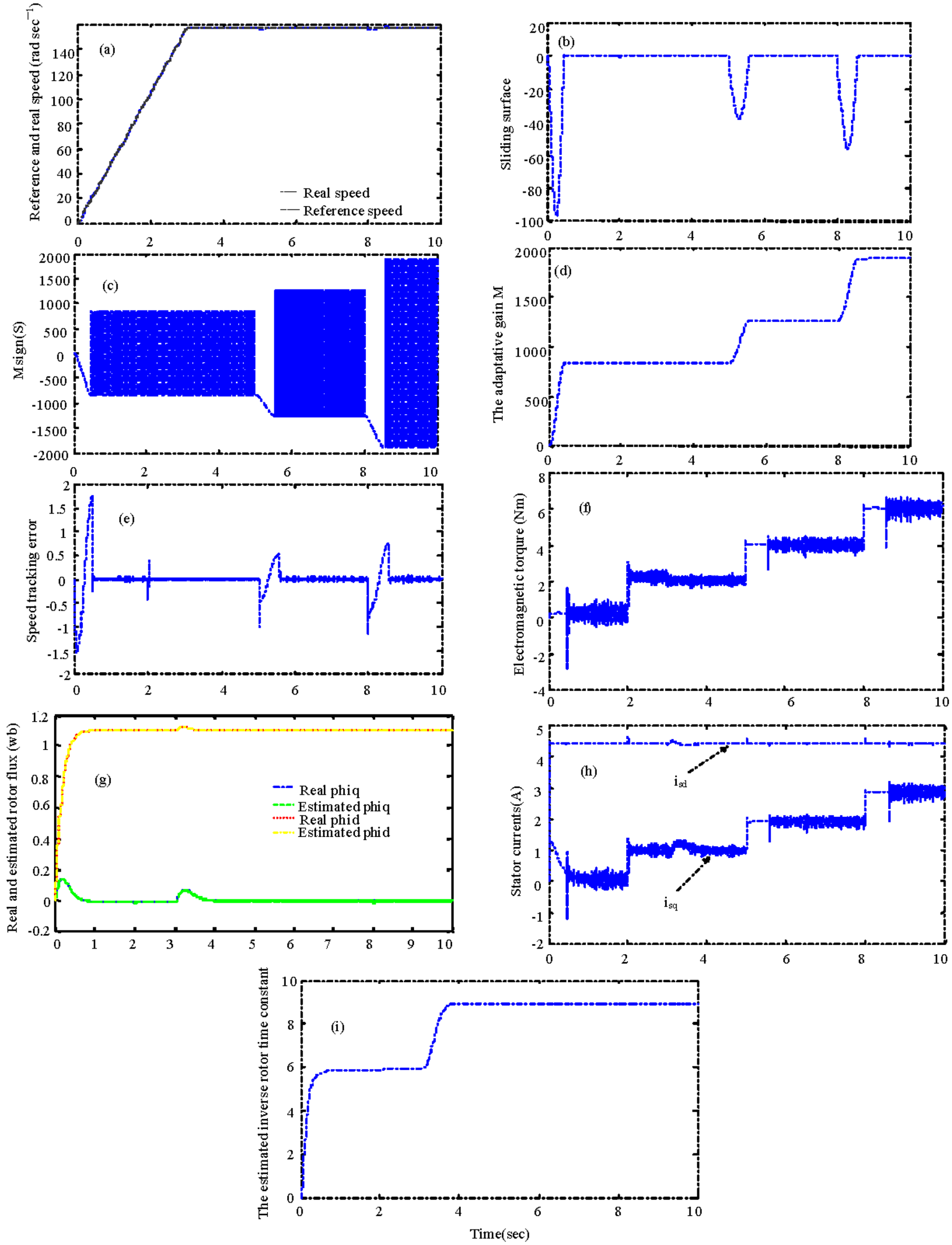

Fig. 4: (a) Real and reference speed (rad/s) (b) Sliding surface (c) Msign(S) (d) The adaptative gain (e) Speed tracking error (f) Electromagnetic torque (Nm) (g) Real and estimated rotor flux (wb) (h) Stator current (A) (i) The estimated inverse rotor time constant

When the sliding surface shown in Fig. 3b converges to zero, the real speed leads the reference one (Fig. 3a) with null tracking error (Fig. 3e) and the adaptative sliding gain converges to a constant value (Fig. 3d). The discontinuous command (Msing(S)) shown in Fig. 3c and 4c, presents a high frequency switching. Figure 3g and 4g show the real and estimated rotor flux components and the decoupling of the machine where the direct flux component leads to reference value and the quadratic one converge to zero. The robustness of the SM observer under rotor resistance variation is clear. Figure 3f and h and 4f and h show the electromagnetic torque and stator currents, respectively. Figure (3i and 4i) show the estimated value of the inverse of the rotor time constant.

When the applied load torque is introduced, the sliding surface becomes different to zero and then no sliding mode exists. In fact, the adaptative sliding gain increases to compensate for the load torque until the sliding surface return to zero. Then the sliding mode occurs again, therefore the sliding gain remains constant. The robustness of the proposed law against rotor resistance variation is proved from Fig. 3 and 4.

CONCLUSION

In this research, an adaptative variable structure control of a DFOC induction motor has been developed. The load torque is considered as an unknown external disturbance that will be rejected by the proposed control law. To achieve the DFOC strategy, an accurate flux observation is needed. Then, the rotor flux is observed using a robust sliding mode observer. The performance of the DFOC technique is ensured by online tuning using the proposed MRAS rotor time constant estimator algorithm. The robustness of the control law and the observer against rotor resistance and load variations is shown via simulation results and the stability is demonstrated using Lyapunov theory. Present research is a part of a SCIM servo drives for robot application. The experimental testes are under construction for a follow up research.

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