

Sensorless Speed Integral Sliding Mode Control with Adaptive Sliding Mode Observer Design of Induction Motor

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Abstract: Sliding-mode control has such advantages as robustness, simple algorithm and good dynamic performance, in this paper we introduce the integral sliding mode control for induction motor based to proposed an integral term in switching surface to have a robust control to parametric and perturbation errors. An accurate knowledge of the rotor speed, rotor flux and adaptatin of resistive parameters are the keys factors in obtaining a high-performance and high-efficiency induction-motor drive, for that,a sliding mode observer design is presented. Simulation results are included to illustrate the performance of sensorless induction motors with flux observer and good results are obtained.

Keywords: Induction motors, integral sliding mode, Sliding mode observer, Sensorless control, Adaptation, Estimation.

1. INTRODUCTION

The new industrial applications necessitate position and speed variators having high dynamic performances, good precision in permanent regime, and a high capacity of overload over the whole range of position and speed, and robustness to different perturbations. Thus, the recourse to robust control algorithms is desirable in stabilization and in tracking trajectories. The variable structure control (VSC) possesses can offer many good properties such as good performance against unmodelled dynamics, insensitivity to parameter variation, external disturbance rejection and fast dynamics [1–4].

Two main approaches, Lyapunov control and Variable Structure Control (VSC), particularly with a sliding mode, have been implemented to provide certain robust stability margin against bounded uncertainty. The advantages of the VSC technique are well known. First, this method enables decomposition of the design problem into two independent stages: selection of discontinuity surfaces with the desired sliding motion, and design of discontinuous control to force the sliding mode

along this manifold. Second, by proper design of a discontinuous controller, the effect of the matched nonlinearities and plant parameter uncertainties and disturbances can be suppressed, and total invariance is obtained when the motion of the system is in sliding mode [5].

The basic principle of sliding mode control consists in moving the state trajectory of the system toward a predetermined surface called sliding or switching surface and in maintaining it around this latter with an appropriate switching logic. This is similar to a feed-forward controller that provides the control that should be applied to track a desired trajectory, which is in this case, the user-defined sliding surface itself. So, the design of a sliding mode controller has two steps, namely, the definition of the adequate switching surface $S(\cdot)$ and the development of the control law or the switching logic U . Concerning the development of the switching logic, it is divided into two parts, the equivalent control U_{eq} and the attractivity or reach ability control U_n . The equivalent control is determined off-line with a model that represents the plant as accurately as possible. If the plant is exactly identical to the model used for determining U_{eq} and there are no disturbances, there would be no need to apply an additional control U_n . However, in practice there will be discrepancy between the model and the actual system control. Therefore, the control component U_n is necessary and it will always guarantee that the state is attracted to the switching surface by satisfying the condition $S(\cdot). \dot{S}(\cdot) < 0$, [6-10].

An observer will be used to construct an estimate of the unmeasured flux states. Several techniques in the literature have been used for flux, speed and (or) parameter estimation for the induction motor. The authors proposed an extended Kalman filter to estimate the rotor flux (or rotor currents) together with the rotor speed and the rotor speed time constant (or rotor resistance), this technique however

is not robust against external disturbances(for example load torque).In other reference the authors used the induction motor equations to estimate the flux. Using two independent subsystems for the rotor flux calculation, an estimation of the rotor speed (considered constant) was given using the model reference adaptive system (MRAS) technique; under load at low frequency this method gives poor results. A linear observer was proposed to estimate the rotor flux when the speed is constant; this technique is not robust against motor parameter variations and requires an adaptation mechanism for parameter identification. In this section we propose a sliding mode rotor flux observer to minimise the resistive parameters effects [11].

In other research, some simple open loop methods can be used to determine the estimated speed on fast way; however they might be sensitive to improper parameters. On the other hand, some closed loop methods using speed observer are robust to mismatched parameters [12].Adaptation of resistive parameters gives object of many works for his important effect; to obtain good performance we use a sliding mode observer.

We can organize this paper as follows; modeling of the induction motor is reviewed in section 2. The different steps for conception of integral sliding mode control are presented in section 3; the sliding mode observer is discussed in section 4. Following by speed and flux observer, adaptive sliding mode observer for resistive parameters deals with the simulation results. Finally some concluding remarks end the paper.

2. INDUCTION MOTOR MODEL

The motor model, under the assumption of linear magnetic circuits, in order fourth is given by:

$$\dot{X} = [A][X] + [B][U] \quad (1)$$

The state vector is X, the control vector (state voltage) is U, where $i_{s\alpha}$ et $i_{s\beta}$ are the stator currents, $\phi_{r\alpha}$ et $\phi_{r\beta}$ are the rotor flux.

$$\begin{cases} [X] = [i_{s\alpha} & i_{s\beta} & \phi_{r\alpha} & \phi_{r\beta}]^T \\ [U] = [V_{s\alpha} & V_{s\beta}]^T \end{cases} \quad (2)$$

For the referential expressed in $(\alpha - \beta)$ stationary reference frame, A and B matrixes are given by:

$$A = \begin{bmatrix} -\frac{R_s}{\sigma L_s} & 0 & \frac{M}{\sigma L_s L_r T_r} & \frac{M}{\sigma L_s L_r} \omega \\ 0 & -\frac{R_s}{\sigma L_s} & -\frac{M}{\sigma L_s L_r} \omega & \frac{M}{\sigma L_s L_r T_r} \\ \frac{M}{T_r} & 0 & -\frac{1}{T_r} & -\omega \\ 0 & \frac{M}{T_r} & \omega & -\frac{1}{T_r} \end{bmatrix} \quad (3)$$

$$B = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \quad (4)$$

3. INTEGRAL SLIDING MODE CONTROL

For a rotor-flux orientation, the regulator imposes the orientation of the rotor flux with respect to the d - axis, giving $\phi_r = \phi_{rd}$ and $\phi_{rq} = 0$. Model motor is given by the following equation system:

$$\begin{aligned} V_{sd} &= \sigma L_s \frac{di_{sd}}{dt} + R_s i_{sd} - \sigma L_s \omega_s i_{sq} - \frac{M}{T_r} \omega_r \phi_r \\ V_{sq} &= \sigma L_s \frac{di_{sq}}{dt} + R_s i_{sq} + \sigma L_s \omega_s i_{sd} - \frac{M}{T_r} \omega_r \phi_r \\ \frac{d\phi_r}{dt} + \frac{1}{T_r} \phi_r &= \frac{M}{T_r} i_{sd} \\ \omega_{sl}^* &= \frac{M i_{sq}^*}{T_r \phi_r} \\ \frac{d\Omega}{dt} &= -\frac{3PM}{2JL_r} (\phi_r i_{sq}^* - C_r - f_r \Omega) \end{aligned} \quad (5)$$

With: $\omega_{sl} = \omega_s - \omega_r$

The proposed control scheme is a cascade structure as it is shown in figure 1, in which four surfaces are required. The internal loops allow the control of the stator current components, whereas the external loops provide the regulation of the speed Ω and the rotor flux ϕ_r .

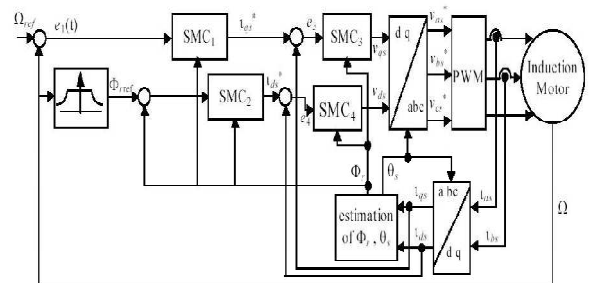


Fig. 1. Block diagram of the integral sliding control

3.1 Design of the switching surfaces

In this work, four sliding surfaces are used and taken as follows since a first order model is used:

$$\begin{aligned} s(\Omega) &= (\Omega_{ref} - \Omega) + m_1 \int (\Omega_{ref} - \Omega) \\ s(\phi_r) &= (\phi_{ref} - \phi_r) + m_2 \int (\phi_{ref} - \phi_r) dt \\ s(i_{sd}) &= (i_{sd}^* - i_{sd}) + m_3 \int (i_{sd}^* - i_{sd}) dt \\ s(i_{sq}) &= (i_{sq}^* - i_{sq}) + m_4 \int (i_{sq}^* - i_{sq}) dt \end{aligned} \quad (6)$$

With Ω_{ref} and ϕ_{ref} being respectively the reference variables of the rotor speed and the flux. $s(\Omega)$, $s(\phi_r)$ are related to the outer loops, whereas $s(i_{sd})$ and $s(i_{sq})$ are related to the inner loops. The i_{sd}^* and i_{sq}^* references are determined by the outer loops, and take respectively the values of the control variables i_{sd} and i_{sq}

3.2 Development of the control laws

We use attractivity condition of switched surface $S(x)\dot{S}(x) < 0$. The vector of control laws can be expressed as:

$$U = U_{eq} + U_n \quad (7)$$

3.2.1 Speed Regulation

By using the equation systems (6) and (7), the regulators control laws are obtained as follows:

$$s(\Omega)\dot{s}(\Omega) < 0 \Rightarrow i_{sq}^* = i_{sq_{eq}} + i_{sq_n} \quad (8)$$

With i_{sq}^* define virtual control input. Decomposed as equivalent component $i_{sq_{eq}}$ and discontinue component i_{sq_n}

A dynamic of switched surface give it as follow:

$$\dot{s}(\Omega) = (\dot{\Omega}_{ref} - \dot{\Omega}) + m_1(\Omega_{ref} - \Omega) \quad (9)$$

By using invariance conditions ($\dot{s} = 0$), equivalent component $i_{sq_{eq}}$ can be expressed as:

$$i_{sq_{eq}} = \frac{2Lr}{3pM\phi_r} \left(J\dot{\Omega}_{ref} + C_r + m_1(\Omega_{ref} - \Omega) \right) \quad (10)$$

And discontinue component expressed as follow:

$$i_{sq_n} = k_1 \text{signs}(\Omega) \quad (11)$$

With: $k_1 > 0$

3.2.2 Rotor flux regulation:

Similar to the first case, we give the same steps for flux regulation as follows:

$$s(\phi_r)\dot{s}(\phi_r) < 0 \Rightarrow i_{sd}^* = i_{sd_{eq}} + i_{sd_n} \quad (12)$$

$$\dot{s}(\phi_r) = (\dot{\phi}_{ref} - \dot{\phi}_r) + m_2(\phi_{ref} - \phi_r)$$

The outer loop of stator current can be expressed as:

$$i_{sd_{eq}} = \frac{T_r}{M} \left(\dot{\phi}_{ref} + \frac{1}{M}\phi_r + m_2(\phi_{ref} - \phi_r) \right) \quad (13)$$

$$i_{sd_n} = k_2 \text{signs}(\phi_r)$$

3.2.3 Stator currents regulation

For the control variables i_{ds} and i_{qs} , of the internal loops are given by:

Regulation of i_{sd}

$$s(i_{sd})\dot{s}(i_{sd}) < 0 \Rightarrow V_{sd} = V_{sd_{eq}} + V_{sd_n} \quad (14)$$

The application of $\dot{s}(i_{sd}) = 0$ we find:

$$\begin{aligned} V_{sd_{eq}} &= \sigma L_s \frac{di_{sd}^*}{dt} + R_s i_{sd}^* - \sigma L_s \omega_s i_{sq} - \frac{M}{T_r} \omega_r \phi_r \\ V_{sd_n} &= k_3 \text{signs}(i_{sd}) \end{aligned} \quad (15)$$

Regulation of i_{sq}

$$s(i_{sq})\dot{s}(i_{sq}) < 0 \Rightarrow V_{sq} = V_{sq_{eq}} + V_{sq_n} \quad (16)$$

With application of invariance condition $\dot{s}(i_{sq}) = 0$ we find:

$$\begin{aligned} V_{sq_{eq}} &= \sigma L_s \frac{di_{sq}^*}{dt} + R_s i_{sq}^* + \sigma L_s \omega_s - \frac{M}{T_r} \omega_r \phi_r \\ V_{sq_n} &= k_4 \text{signs}(i_{sq}) \end{aligned} \quad (17)$$

To satisfy the stability condition of the system, the gains K_1, K_2, K_3 and K_4 should be taken positive by selecting the appropriate values.

4. SLIDING MODE OBSERVER

Consider only the first four equations of the induction motor model given by equation (1). In the following exposition, the speed will be considered as a varying parameter. The proposed observer aims is to estimate firstly the rotor flux. The observer is given by the following system:

$$\frac{d}{dt} \begin{bmatrix} \hat{i}_s \\ \hat{\phi}_r \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{\sigma L_s} I & \frac{M}{\sigma L_s L_r T_r} I - j \frac{\omega}{\sigma L_s} \\ \frac{M^2}{L_r T_r} I & -\frac{M}{L_r T_r} I + j \omega \end{bmatrix} \begin{bmatrix} i_s \\ \phi_r \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \end{bmatrix} U_s + \begin{bmatrix} K_s \\ K_r \end{bmatrix} \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} \quad (18)$$

Where:

$$I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad j = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}, \quad K = \begin{bmatrix} K_s \\ K_r \end{bmatrix}.$$

$S = \begin{bmatrix} S_1 \\ S_2 \end{bmatrix} = m \begin{bmatrix} i_{s\alpha} - \hat{i}_{s\alpha} \\ i_{s\beta} - \hat{i}_{s\beta} \end{bmatrix}$ Is the sliding surface which

represents the error between the measured current components and those estimated.

Thus, we use the K_r gain for action to fix the dynamics of convergence of the evaluation error flux (reduced system is equivalent) [6]. One considers then, the error of estimation of the variables of states is given by:

$$\begin{bmatrix} \dot{\hat{i}}_s \\ \dot{\hat{\phi}}_r \end{bmatrix} = \begin{bmatrix} 0 & \frac{M}{\sigma L_s L_r T_r} I - j \frac{\omega}{\sigma L_s} \\ 0 & -\frac{M}{L_r T_r} I + j \omega \end{bmatrix} \begin{bmatrix} \epsilon_{is} \\ \epsilon_{\phi} \end{bmatrix} - \begin{bmatrix} K_s \\ K_r \end{bmatrix} \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} - \Delta A_{R_s} \begin{bmatrix} i_s \\ \hat{\phi}_r \end{bmatrix} - \Delta A_{R_r} \begin{bmatrix} i_s \\ \hat{\phi}_r \end{bmatrix} \quad (19)$$

Which ΔR_s and ΔR_r represent the modelling error of the resistive parameters R_s and R_r , then:

$$\Delta A_{R_s} = \begin{bmatrix} \frac{R_s - \hat{R}_s}{\sigma L_s} I & 0 \\ 0 & 0 \end{bmatrix} \quad (20)$$

$$\Delta A_{R_r} = \begin{bmatrix} M^2 \frac{R_r - \hat{R}_r}{\sigma L_s L_r^2} I & -M^2 \frac{R_r - \hat{R}_r}{\sigma L_s L_r^2} I \\ -M^2 \frac{R_r - \hat{R}_r}{L_r^2} I & M^2 \frac{R_r - \hat{R}_r}{L_r^2} I \end{bmatrix} \quad (21)$$

4.1 Matrix gains correction Ks

In a first time, when we don't consider the modelling error, we consider that is $\Delta R_s = 0$ and $\Delta R_r = 0$, then the equation of the error of estimation of the limits states itself to:

$$\begin{bmatrix} \dot{\epsilon}_{is} \\ \dot{\epsilon}_{\phi} \end{bmatrix} = \begin{bmatrix} 0 & \frac{M}{\sigma L_s L_r T_r} I - j \frac{\omega}{\sigma L_s} \\ 0 & -\frac{M}{L_r T_r} I + j \omega \end{bmatrix} \begin{bmatrix} \epsilon_{is} \\ \epsilon_{\phi} \end{bmatrix} - \begin{bmatrix} K_s \\ K_r \end{bmatrix} \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} \quad (22)$$

To assure the asymptotic convergence of S toward zero, one search for the necessary conditions of stability, bound to the values of the K_s gain. While choosing the following Lyapunov function:

$$V = \frac{1}{2} S^T S \quad (23)$$

In order to assure the convergence S to zero, we must verify that the derivative of V is strictly negative because V is a positive function.

$$\dot{V} = S^T \dot{S} < 0 \Rightarrow \dot{V} = S^T m \dot{\epsilon}_s < 0 \quad (24)$$

$$\dot{V} = [S_1 \ S_2] \begin{bmatrix} 1 & \alpha \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \frac{1}{\sigma L_s} \left(\frac{M}{L_r T_r} I - j \omega \right) \epsilon_{\phi} - K_s \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} \end{bmatrix} < 0 \quad (25)$$

If one puts:

$$K_s = m^{-1} \begin{bmatrix} G_1 & 0 \\ 0 & G_2 \end{bmatrix} \quad (26)$$

Then the derivative of the Lyapunov function becomes negative:

$$\dot{V} = S_1 (f_1 + \alpha f_2 - G_1 \text{sgn}(S_1)) + S_2 (f_2 - G_2 \text{sgn}(S_2)) < 0 \quad (27)$$

To assure the convergence of S toward zero one must verify the following conditions:

If $S_1 > 0$ then $G_1 > f_1 + \alpha f_2$; else $S_1 < 0$ alors $-G_1 < f_1 + \alpha f_2$

If $S_2 > 0$ then $G_2 > f_2$; else $S_2 < 0$ then $-G_2 < f_2$

So $G_1 > |f_1 + \alpha f_2|$, and $G_2 > |f_2|$.

It only remains to choose values of G_1 and sufficiently big G_2 for to verify the convergence of S

toward zero, that is ($\hat{i}_s \rightarrow i_s$).

4.2 Matrix gains correction Kr

In order to calculate the gain Kr of correction of the equivalent scale model, we consider that we slide on the surface S. ($S = 0, \dot{S} = 0$), this hypothesis is verified, when the dynamic of the stator currents is rapid; then while considering that the current error considering as follows:

Current's error $\varepsilon_{is} = 0, \dot{\varepsilon}_{is} = 0$

We will have then:

$$\dot{\varepsilon}_{is} = \frac{1}{\sigma L_s} \left[\frac{M}{L_r T_r} I - j\omega \right] \varepsilon_{\phi} - K_s \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} = 0 \quad (28)$$

Then we have sliding surface given by (28).

$$\begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} = \frac{1}{\sigma L_s} K_s^{-1} \left[\frac{M}{L_r T_r} I - j\omega \right] \varepsilon_{\phi} \quad (29)$$

With this last expression, we can express the equation of the error of the rotor flux by:

$$\dot{\varepsilon}_{\phi} = \left(-\frac{M}{L_r T_r} I + j\omega - K_r K_s^{-1} \left[\frac{M}{L_r T_r} I - j\omega \right] \right) \varepsilon_{\phi} \quad (30)$$

We calculate the gain of Kr correction by identification to an equivalent system with dynamics that assures the desire behaviour

$$\dot{\varepsilon}_{\phi} = -Q \varepsilon_{\phi} \quad (31)$$

$$Q = \begin{bmatrix} q_1 & 0 \\ 0 & q_2 \end{bmatrix} \quad (32)$$

Q: Matrix gain defined positive.

We can recover the expression of Kr by:

$$K_r = \frac{\sigma L_s}{\beta} \begin{bmatrix} G_1 \left(\frac{M}{L_r T_r} q_1 - \beta \right) & -G_2 \left(q_1 \omega + \alpha \left(\frac{M}{L_r T_r} q_1 - \beta \right) \right) \\ G_1 \omega q_2 & G_2 \left(\frac{M}{L_r T_r} q_2 - \beta - \alpha \omega q_2 \right) \end{bmatrix} \quad (33)$$

Where:

$$\beta = \left(\frac{M}{L_r T_r} \right)^2 + \omega^2$$

4.3 Speed observer design

We consider the perfect motor model motor

(supposing that resistances and inductances are perfectly known) and also the speed supposed as a parameter when incertitude exists and is presented by:

$$\hat{\omega} = \omega + \Delta \omega \quad (34)$$

$$\begin{bmatrix} \dot{\varepsilon}_{is} \\ \dot{\varepsilon}_{\phi} \end{bmatrix} = \begin{bmatrix} 0 & \left(\frac{M}{\sigma L_s L_r T_r} \right) I - j\omega \\ 0 & -\left(\frac{M}{L_r T_r} \right) I + j\omega \end{bmatrix} \begin{bmatrix} \varepsilon_{is} \\ \varepsilon_{\phi} \end{bmatrix} - \begin{bmatrix} K_s \\ K_r \end{bmatrix} \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} - \Delta A_{\omega}^t \begin{bmatrix} i_s \\ \hat{\phi}_r \end{bmatrix} \quad (35)$$

Where:

$$\Delta A_{\omega}^t = (\omega - \hat{\omega}) \begin{bmatrix} 0 & -\frac{J}{\sigma L_s} \\ 0 & J \end{bmatrix}$$

In this condition we can develop the derivation of V as:

$$\dot{V} = S_1(f_1 + \alpha f_2 - G_1 \text{sgn}(S_1)) + S_2(f_2 - G_2 \text{sgn}(S_2)) - S^t m(\omega - \hat{\omega}) \frac{J}{\sigma L_s} \hat{\phi}_r \quad (36)$$

To compensate this term $S^t m(\omega - \hat{\omega}) \frac{J}{\sigma L_s} \hat{\phi}_r$ (we don't know his sign) the Lyapunov equation considerate as:

$$V_2 = V + \frac{(\omega - \hat{\omega})^2}{2\lambda_3} \quad (37)$$

When: $\lambda_3 > 0$

The derivation of V_2 is presented in equation (38):

$$\dot{V}_2 = S_1(f_1 + \alpha f_2 - G_1 \text{sgn}(S_1)) + S_2(f_2 - G_2 \text{sgn}(S_2)) - S^t m(\omega - \hat{\omega}) \frac{J}{\sigma L_s} \hat{\phi}_r - \frac{\omega - \hat{\omega}}{\lambda_3} \hat{\omega} \quad (38)$$

To assure $\dot{V}_2 < 0$ for the development of adaptation low of speed we need to propose:

$$\dot{\hat{\omega}} = -\lambda_3 \frac{S^t m J \hat{\phi}_r}{\sigma L_s} \quad (39)$$

4.4 Resistive parameters Adaptation

We use estimation error, and we add modeling errors caused by resistive parameters, for this the derivative of Lyapunov function chosen as follow:

$$\begin{aligned} \dot{V} = S^t m \left(\frac{1}{\sigma L_s} \left(\frac{M}{L_r T_r} I - j\omega \right) \varepsilon_{\phi_r} - K_s \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} \right) \\ - S^t m \frac{R_s - \hat{R}_s}{\sigma L_s} i_s - S^t m M^2 \frac{R_r - \hat{R}_r}{\sigma L_s L_r^2} I \left(i_s - \frac{\hat{\phi}_r}{M} \right) \end{aligned} \quad (40)$$

To guarantee Lyapunov condition $\dot{V} < 0$, we take this two equation mentioned in (41) to zero because we don't know his sign:

$$S^t m \frac{R_s - \hat{R}_s}{\sigma L_s} i_s; \text{ and } S^t m M^2 \frac{R_r - \hat{R}_r}{\sigma L_s L_r^2} I \left(i_s - \frac{\hat{\phi}_r}{M} \right) \quad (41)$$

After compensation of these terms we propose another function with his derivative as:

$$V_1 = \frac{1}{2} S^t S + \frac{(R_s - \hat{R}_s)^2}{2\lambda_1} + \frac{(R_r - \hat{R}_r)^2}{2\lambda_2} \quad (42)$$

$$\begin{aligned} \dot{V}_1 = S_1(f_1 + \alpha_2 - G_1 \text{sgn}(S_1)) + S_2(f_2 - G_2 \text{sgn}(S_2)) - S^t m \frac{R_s - \hat{R}_s}{\sigma L_s} i_s \\ - S^t m M^2 \frac{R_r - \hat{R}_r}{\sigma L_s L_r^2} I \left(i_s - \frac{\hat{\phi}_r}{M} \right) - \frac{R_s - \hat{R}_s}{\lambda_1} - \frac{R_r - \hat{R}_r}{\lambda_2} \end{aligned} \quad (43)$$

In the end we present estimated resistive parameters as follow:

$$\hat{R}_s = -\lambda_1 S^t \frac{M}{\sigma L_s} i_s \quad (44)$$

$$\hat{R}_r = -\lambda_2 S^t \frac{M}{\sigma L_s} \left(i_s - \frac{\hat{\phi}_r}{M} \right) \quad (45)$$

Where: $\lambda_1, \lambda_2 > 0$

5. SIMULATION RESULTS

The effectiveness of the proposed controller combined with the rotor flux and rotor speed estimation has been verified by simulations. The simulation results have been obtained under a constant load of 5 N.m at 0.5 s.

Table 1. Motor Parameters

R_s	10 Ω	L_s	0.4642 H
R_r	6.3 Ω	L_r	0.4612 H
M	0.4212 H	J	0.02 kgm^2

AND: P=7.5 kW, 220 V, 50 Hz.

The speed tracking controller is operated in a critical situation (rapidly changes as 157, -157, 0, 5 rad/s). Figure 1 show the satisfactory performances of the speed tracking. We can see that the actual speed follows the speed command and estimated speed. Thus, the simulation results confirm that the proposed observer give good results justified by rotor speed error converges to zero rapidly

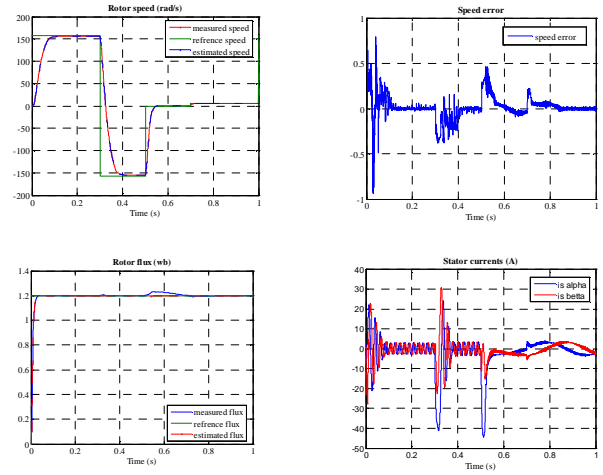


Fig. 2. Speed variation of integral sliding control with sliding mode observer

In This time we applied variation of 100% of the nominal rotor resistance between time $t = 0.3\text{s}$ and $t=0.7\text{s}$. Figure (3) presents the obtained results. We can mention good observation for speed and flux, the effect of rotor resistance variation on the speed is negligible. And error speed is converged to zero.

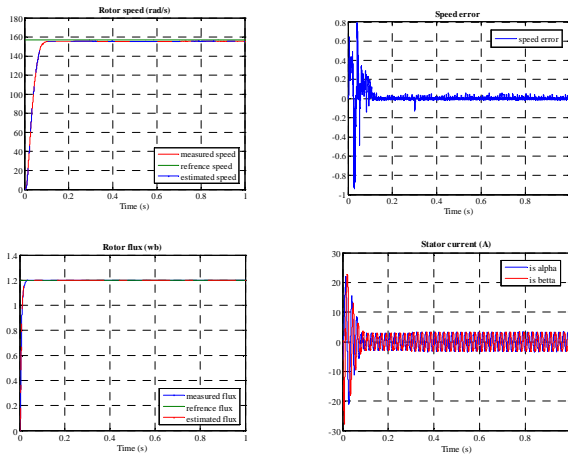


Fig. 3. Rotor resistance variation of sliding mode control with sliding mode observer

In this last figure we combine all observers (flux, speed and adaptation of resistive parameters). We obtain good results for our sliding mode observer with integral sliding mode control in nominal conditions.

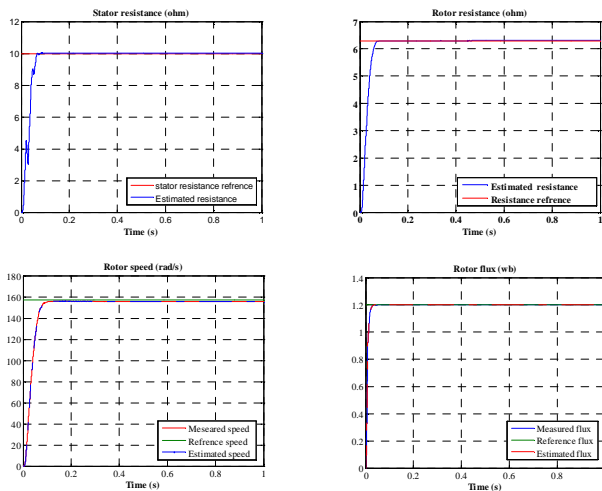


Fig.4 Elaboration of observers with integral sliding control

7. CONCLUSION

In this work we have presented integral sliding mode control in order to offer a choice of design tools to accommodate uncertainties and nonlinearities. This study has successfully demonstrated the design of the sliding mode technique of an induction motor. In this paper we made also the conception of rotor flux observer and with sensorless speed control with adaptation of resistive parameters, a good precision;

it is proved, with simulation results obtained with the different tests of robustness executed. These good results encourage us to say that the observer is robust.

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LIST OF SYMBOLS

IM	Induction motor
s, r	Indices for stator and rotor.
R_s, R_r	Stator and rotor resistance.
R_t	Total resistance referred to the stator
L_s, L_r	Self inductance of stator and rotor
M	Mutual inductance.
J	Inertia moment of the moving element
F	Viscous friction coefficient.
T_r, T_s	Rotor and stator time constant.
σ	Is the coefficient of dispersion
ϕ_s, ϕ_r	Stator and rotor flux.
C_{em}	Electromagnetic torque.
C_r	Load torque.
ω_s, ω_r	Stator and rotor angular frequency .
P	Is number of pole pairs
θ_r	Angular position of the moving element.
Θ	Angular position.
$\phi_{r\text{ref}}$	Rotor flux reference.
(d, q)	Axes for direct and quadrature park subscripts.
(α, β)	Axes for stationary reference frame subscripts
$V(x)$	Lyapunov Function.
$S(x)$	Sliding surface.
J	Imaginary matrix
I	Identity matrix
$\hat{}$	Estimated sign