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INDUCTION MOTOR CONTROL BASED ON A FUZZY SLIDING MODE APPROACH

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Key words: Induction motor (IM), Non-linear control, Speed regulation, Fuzzy logic control (FLC), Sliding mode control (SMC), Fuzzy sliding mode control (FSMC), Chattering phenomena, Parameters variation, Robustness.

This paper presents the design of a hybrid fuzzy sliding mode controller for a three-phase induction motor controlled by the technique of rotor flux orientation. The proposed approach combines sliding mode control with the fuzzy logic control to accurately track predefined trajectories, ensuring robust performance despite parameter variations and load disturbances. The combination of the two approaches reduces the high-frequency oscillations caused by the chattering phenomenon, without affecting the accuracy and robustness of the control. The design of the hybrid control law is described and numerical simulation results are presented and compared with the performance of conventional sliding mode control. These results clearly show the reduction of the chattering and the effectiveness of the proposed method in different operating conditions.

1. INTRODUCTION

Induction motor is the most widely used electric actuator in the industrial field. Indeed, it is highly appreciated for its several advantages such as its low cost, its high robustness, its high mass torque, its low inertia and absence of a brush-collector system [1]. It is very attractive for wide areas of application and in difficult environments, for example in the field of machine tools, pumping and ventilation systems, handling equipment or high-performance electric drives such as traction electrical applications.

In recent years, there have been many advances in variable speed applications using induction motors. Today, the vector control with rotor flux oriented (RFOC) has become the most used strategy, allowing to independently the control of the flux and the electromagnetic torque. This technique can provide a high level of performance; however, it requires perfect knowledge of the parameters of the considered system, which is not the case for electrical machines. On the other hand, linear controllers, such as conventional PI and proportional integral derivative (PID) controllers used in vector control, have insufficient performance and do not make it possible to obtain accurate and robust responses for non-linear systems. Consequently, several non-linear techniques have been developed to improve performances with an increase to robustness of the control against the parametric variations and external disturbances. Among the control strategies proposed to solve these problems are exact linearization techniques [2], sliding mode control (SMC) [3], and backstepping control [4], the approaches based on the passivity [5] and flatness [6], the intelligent controls such as artificial neural networks (ANN) [7] or fuzzy logic control (FLC) [8].

The non-linear sliding mode approach has been successfully applied to control the induction machine. The influence from the variability of some parameters, especially the rotor time constant, and the sensitivity to load disturbances are thus greatly reduced. This control makes it possible to design variable speed drives for induction motor, with high performance in transient and steady state conditions, characterized by high robustness [9].

However, the major disadvantage of this control is the phenomenon of chattering, which creates high frequency oscillations due to the discontinuous properties of this technique. These oscillations can cause early wears and damage to the systems by exciting neglected dynamics and eventually degrade the actuators and equipment [3–10].

To reduce the effects of chattering, some authors have developed alternatives to the classic SMC. Some solutions use a smoothed continuous switching function or a thin layer defined around the sliding surface limiting the switching within it [10]. However, a compromise is necessary between the chattering limitation and the accuracy of tracking control which deteriorates with the limitation of available bandwidth [11]. The super twisting method has been widely used for position and speed control of ac machines. It is a second-order sliding mode algorithm that improves performances and reduces chattering oscillations, but has a high degree of complexity and its adjustment technique remains difficult [11].

Recently, researchers have applied several strategies, based on hybrid fuzzy sliding mode control (FSMC) for the IM. The aim is to obtain a better compromise between weak chattering oscillations and good tracking accuracy under uncertain load disturbances. The combination of the two approaches is motivated by the similarities between SMC and FLC techniques [12–14]. These methods involve replacing the switching function by a fuzzy inference system (FIS) to increase the boundary layer around the switching surface with smoother switching.

In this paper, a sliding mode control law is presented, the basic idea consists, retaining a classical switching function, to design an equivalent control based on fuzzy logic and thus obtains a fast and robust dynamic, limiting the chattering by the combined actions of the two approaches.

This paper is organized as follows: Section 2 is devoted to the IM modeling in the (*d-q*) rotating frame, and then the principle of the orientation of the rotor flux is presented. Sections 3 shows first, the classical theory of SMC, then deal with the design of the FSMC proposed. Simulation tests and results are presented and discussed in Section 4. Finally, Section 5 ends the paper with a conclusion.

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2. **FIELD ORIENTED CONTROL**

2.1 MODEL OF IM IN (*d-q*) REFERENCE FRAME

The induction motor can be modeled, using the Park transformation, in (*d-q*) synchronously rotating frame, with an appropriately selecting the stator currents and rotor flux as state variables, by the following non-linear equations [1]:

$$\begin{cases} \dfrac{d i_{sd}}{dt} = -\gamma i_{sd} + \omega_s i_{sq} + \dfrac{K}{\tau_r}\phi_{rd} + K\omega_r\phi_{rq} + \dfrac{v_{sd}}{\sigma L_s} \\ \dfrac{d i_{sq}}{dt} = -\omega_s i_{sd} - \gamma i_{sq} - K\omega_r\phi_{rd} + \dfrac{K}{\tau_r}\phi_{rq} + \dfrac{v_{sq}}{\sigma L_s} \\ \dfrac{d\phi_{rd}}{dt} = \dfrac{L_m}{\tau_r} i_{sd} - \dfrac{1}{\tau_r}\phi_{rd} + (\omega_s - \omega_r)\phi_{rq} \\ \dfrac{d\phi_{rq}}{dt} = \dfrac{L_m}{\tau_r} i_{sq} - (\omega_s - \omega_r)\phi_{rd} - \dfrac{1}{\tau_r}\phi_{rq} \\ \dfrac{d\omega_r}{dt} = K_T(\phi_{rd} i_{sq} - \phi_{rq} i_{sd}) - \dfrac{p}{J}T_l - \dfrac{B}{J}\omega_r \end{cases}, \quad (1)$$

with

$$\gamma = \frac{R_s}{\sigma L_s} + \frac{R_r L_m^2}{\sigma L_s L_r^2} \qquad K = \frac{L_m}{\sigma L_s L_r}.$$

$$K_T = \frac{3}{2}\frac{p^2 L_m}{J L_r} \qquad \sigma = 1 - \frac{L_m^2}{L_s L_r} \qquad \tau_r = \frac{L_r}{R_r}.$$

(v_{sd},v_{sq}) and (i_{sd},i_{sq}) respectively represented the direct and quadrature components of the stator voltages and currents. (ϕ_{rd}, ϕ_{rq}) denote the (*d-q*) components of the rotor flux.

p is the pole pair number of the motor; J is the total inertia of the system and B is the viscous damping coefficient. R_s and R_r are the stator and rotor winding resistances; L_s, L_r and L_m are the stator, rotor and mutual inductances.

σ and τ_r are the total leakage factor and rotor time constant. ω_r and ω_{sl} are the rotor and slip frequencies.

The angular synchronous frequency is given by:

$$\omega_s = \omega_r + \omega_{sl}. \quad (2)$$

The mechanical speed of the motor shaft is expressed as:

$$\Omega_m = \frac{\omega_r}{p}. \quad (3)$$

The electromagnetic torque produced by the induction motor is given by the relation:

$$T_{em} = K_T(\phi_{rd} i_{sq} - \phi_{rq} i_{sd}). \quad (4)$$

The load torque T_l is considered an external disturbance.

2.2 ROTOR FLUX ORIENTATION

The aim of the vector control with rotor flux orientation for the IM is to independently control the electromagnetic torque and flux.

The principle consists in aligning the *d* axis of the revolving reference frame with the rotor flux vector ϕ_r, as illustrated in Fig. 1 [1].

Thus, the orientation condition of the rotor flux is:

$$\phi_{rq} = 0 \quad \text{and} \quad \phi_r = \phi_{rd} = \text{constant}. \quad (5)$$

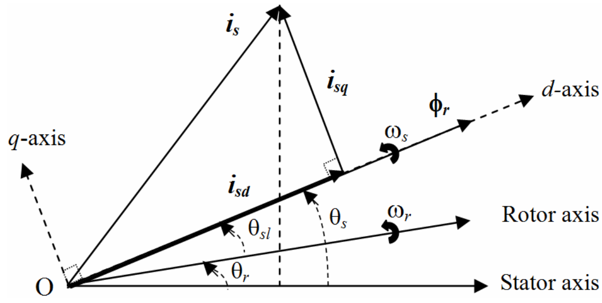

Fig.1 – Angular positions and alignment of the revolving reference frame with the rotor flux vector.

The electromagnetic torque expression becomes:

$$T_{em} = K_T \phi_{rd} i_{sq}. \quad (6)$$

The spatial vectors of the rotor flux ϕ_r and stator quadrature current i_{sq} are orthogonal and thus the torque and the flux are entirely decoupled as for the dc motor.

According to equation (5), the model of the motor can be expressed by the following equations:

$$\begin{cases} \dfrac{d i_{sd}}{dt} = -\gamma i_{sd} + \omega_s i_{sq} + \dfrac{K}{\tau_r}\phi_r + \dfrac{v_{sd}}{\sigma L_s} \\ \dfrac{d i_{sq}}{dt} = -\omega_s i_{sd} - \gamma i_{sq} - K\omega_r\phi_r + \dfrac{v_{sq}}{\sigma L_s} \\ \dfrac{d\phi_r}{dt} = \dfrac{L_m}{\tau_r} i_{sd} - \dfrac{1}{\tau_r}\phi_r \\ 0 = \dfrac{L_m}{\tau_r} i_{sq} - (\omega_s - \omega_r)\phi_r \\ \dfrac{d\omega_r}{dt} = K_T\phi_r i_{sq} - \dfrac{p}{J}T_l - \dfrac{B}{J}\omega_r \end{cases}. \quad (7)$$

In steady state, its value ϕ_r of the flux is as follows:

$$\phi_r = L_m i_{sq}. \quad (8)$$

The rotor flux can be adjusted and maintained constant by the direct current component i_{sd}, while the quadrature current component i_{sq} is used to control the torque [1].

The slip frequency can be calculated from the values of the stator current i_{sq} and the rotor flux reference as follow:

$$\omega_{sl} = \omega_s - \omega_r = \frac{L_m}{\tau_r}\frac{i_{sq}}{\phi_r} \quad (9)$$

and the angular position θ_s of the rotor flux is given by:

$$\theta_s = \int \omega_s \, dt = \int\left(p\Omega + \frac{L_m}{\tau_r}\frac{i_{sq}}{\phi_r}\right)dt. \quad (10)$$

The flux angular position θ_s is used to perform the direct and inverse Park transformations, and thus guarantee a correct alignment of the revolving reference frame.

3. **DESIGN OF FUZZY SLIDING MODE CONTROL**

3.1 SLIDING MODE THEORY

The sliding modes control technique is a particular type of the variable structure control system (VSCS). Its principle consists in forcing the state trajectory of a system towards a sliding surface and to maintain it around in its vicinity, with appropriate switching logic [3–9].

The sliding surface represents the desired equilibrium state, while the state trajectory to reach the surface imposes the transient dynamic response of the system.

Let a non-linear system given by the canonical form:

$$\begin{cases} \dot{\boldsymbol{x}}(t) = f(\boldsymbol{x},t) + g(\boldsymbol{x},t)\boldsymbol{u}(t) + \boldsymbol{d}(t) \\ \boldsymbol{y}(t) = \boldsymbol{x}(t) \end{cases}. \tag{11}$$

$\boldsymbol{x}(t) \in \Re^n$, $\boldsymbol{u}(t) \in \Re^m$, and $\boldsymbol{y}(t) \in \Re^n$ are respectively the state, input and output vectors of the system; $f(\boldsymbol{x},t)$, and $g(\boldsymbol{x},t)$ are non-linear functions; $\boldsymbol{d}(t)$ represents external disturbances assumed to be bounded as $\boldsymbol{d}(t) \leq \boldsymbol{D}(t)$.

The objective is to establish a control law $\boldsymbol{u}(t)$ forcing the output of the system $\boldsymbol{y}(t)$ to accurately follow the reference trajectories $\boldsymbol{y}^*(t)$ in finite time from an arbitrary initial condition, and thus cancel the tracking errors given by:

$$\boldsymbol{e}(t) = \boldsymbol{y}^*(t) - \boldsymbol{y}(t) = \boldsymbol{x}^*(t) - \boldsymbol{x}(t). \tag{12}$$

The general equation, proposed by J.J. Slotine [15], to determine the sliding surfaces and to ensure the convergence of the variables to their desired value is as:

$$\boldsymbol{S}(\boldsymbol{x}) = \left(\frac{\mathrm{d}}{\mathrm{d}t} + \lambda\right)^{r-1} \boldsymbol{e}(\boldsymbol{x}) \qquad (\lambda \succ 0) \tag{13}$$

r is the relative degree, equal to the derivation number of the output necessary to make the command appear.

The system enters in the sliding mode, with $\boldsymbol{S}(\boldsymbol{x})= 0$, if the convergence condition (14) is satisfied [15]:

$$\boldsymbol{S}(\boldsymbol{x})\dot{\boldsymbol{S}}(\boldsymbol{x}) \prec 0 \tag{14}$$

This condition indicates that the sliding surface becomes attractive and invariant, the sliding mode is then ideal with an infinite switching frequency.

The control law is obtained by adding two terms such as:

$$\boldsymbol{u}_{smc} = \boldsymbol{u}_{eq} + \boldsymbol{u}_n \tag{15}$$

$\boldsymbol{u}_{eq}$ is the equivalent control that depicts a low-frequency term that influences the approach mode (reaching mode) to reach the switching surface, $\boldsymbol{u}_n$ is the switching control, it is a discontinuous term of high-frequency that influences the sliding mode by maintaining the state of the system on the switching surface.

The general form of discontinuous control is then [16]:

$$\boldsymbol{u}_n = -\boldsymbol{k}\,\mathrm{sign}[\boldsymbol{S}(x)] \tag{16}$$

$\boldsymbol{k}$ is a diagonal matrix with constant and positive coefficients allows adjusting the desired dynamics.

In fact, the ideal sliding mode does not occur because the control should then change with infinite frequency. This is impossible due to a delay time for the control calculation and the limits of the switching frequency imposed to switches of the inverter. Thus, the state trajectory does not move exactly along the surface, but tends to oscillate around it [3]. This behavior, called chattering, causes high frequency oscillations due to the discontinuous properties of the control.

The chattering is prejudicial to the proper operating; it can deteriorate the system by exciting the neglected dynamics during the modeling, or damage the actuators by too frequent constraints [3–14].

3.2 SLIDING MODE CONTROL FOR IM

The non-linear equations (7) with the orientation of the rotor flux can be rewritten as follows:

$$\begin{cases} \dot{\boldsymbol{x}} = f(\boldsymbol{x}) + g(\boldsymbol{x})\boldsymbol{u} \\ \boldsymbol{y} = \boldsymbol{x} \end{cases}, \tag{17}$$

where $\boldsymbol{x} = [i_{sd}\ \ i_{sq}\ \ \phi_r\ \ \omega_r]^{\mathrm{T}}$ is the state vector, $\boldsymbol{u} = [v_{sd}\ \ v_{sq}]^{\mathrm{T}}$ is the control vector; $f(\boldsymbol{x})$ and $g(\boldsymbol{x})$ are functions defined by:

$$f(\boldsymbol{x}) = \begin{bmatrix} f_1(\boldsymbol{x}) \\ f_2(\boldsymbol{x}) \\ f_3(\boldsymbol{x}) \\ f_4(\boldsymbol{x}) \end{bmatrix} = \begin{bmatrix} -\gamma x_1 + \omega_s x_2 + \dfrac{K}{\tau_r} x_3 \\ -\omega_s x_1 - \gamma x_2 - K x_3 x_4 \\ \dfrac{L_m}{\tau_r} x_1 - \dfrac{1}{\tau_r} x_3 \\ K_T x_2 x_3 - \dfrac{p}{J} T_l - \dfrac{B}{J} x_4 \end{bmatrix}; \quad g(\boldsymbol{x}) = \begin{bmatrix} \dfrac{1}{\sigma L_s} & 0 & 0 & 0 \\ 0 & \dfrac{1}{\sigma L_s} & 0 & 0 \end{bmatrix}^{\mathrm{T}}. \tag{18}$$

Considering the size of the control vector $\boldsymbol{u}$ and knowing that the variables to be adjusted are rotor flux ϕ_r and speed ω_r, the chosen sliding surface is given by:

$$\boldsymbol{S} = \boldsymbol{\lambda e} + \dot{\boldsymbol{e}}\ . \tag{19}$$

with

$$\boldsymbol{S} = \begin{bmatrix} S_\phi \\ S_\omega \end{bmatrix}; \quad \boldsymbol{e} = \begin{bmatrix} e_\phi \\ e_\omega \end{bmatrix} = \begin{bmatrix} \phi_r^* - \phi_r \\ \omega_r^* - \omega_r \end{bmatrix}; \quad \boldsymbol{\lambda} = \begin{bmatrix} \lambda_\phi & 0 \\ 0 & \lambda_\omega \end{bmatrix}.$$

ω_r^* and ϕ_r^* are the flux and the reference speed, λ_ϕ and λ_ω are positive coefficients.

To make the sliding surface invariant ($\dot{\boldsymbol{S}}=0$) and attractive ($\boldsymbol{S}^{\mathrm{T}}\dot{\boldsymbol{S}}<0$), the following dynamics are defined:

$$\dot{\boldsymbol{S}} = \boldsymbol{\lambda}\dot{\boldsymbol{e}} + \ddot{\boldsymbol{e}} = \boldsymbol{Q} + \boldsymbol{R}\boldsymbol{u}\ , \tag{20}$$

where

$$\boldsymbol{Q} = \begin{bmatrix} \ddot{\phi}_r^* + \lambda_\phi \dot{\phi}_r^* - \left(\lambda_\phi - \dfrac{1}{\tau_r}\right) f_3 - \dfrac{L_m}{\tau_r} f_1 \\ \ddot{\omega}_r^* + \lambda_\omega \dot{\omega}_r^* + \dfrac{p}{J}\dot{T}_l - \left(\lambda_\omega - \dfrac{B}{J}\right) f_4 - k_T\left(\phi_r f_2 + i_{sq} f_3\right) \end{bmatrix}.$$

$$\boldsymbol{R} = -\frac{1}{\sigma L_s}\begin{bmatrix} \dfrac{L_m}{\tau_r} & 0 \\ 0 & k_T \phi_r \end{bmatrix}.$$

During the reaching mode, the derivative of $\boldsymbol{S}$ is zero, so:

$$\dot{\boldsymbol{S}} = \boldsymbol{Q} + \boldsymbol{R}\boldsymbol{u}_{eq} = 0\ . \tag{21}$$

and the equivalent controls are then expressed by:

$$\boldsymbol{u}_{eq} = \begin{bmatrix} v_{sd\,eq} \\ v_{sq\,eq} \end{bmatrix} = -\boldsymbol{R}^{-1}\boldsymbol{Q}\ . \tag{22}$$

while in the sliding mode, the control law is:

$$\boldsymbol{u}_{smc} = \begin{bmatrix} v_{sd\,smc} \\ v_{sq\,smc} \end{bmatrix} = \begin{bmatrix} v_{sd\,eq} \\ v_{sq\,eq} \end{bmatrix} + \begin{bmatrix} v_{sd\,n} \\ v_{sq\,n} \end{bmatrix}. \tag{23}$$

To track the trajectories ϕ_r^* and ω_r^*, it is possible to choose the following control law for the stator voltages:

$$\begin{bmatrix} v_{sd\,smc} \\ v_{sq\,smc} \end{bmatrix} = -\boldsymbol{R}^{-1}\boldsymbol{Q} - \boldsymbol{R}^{-1}\begin{bmatrix} k_\phi & 0 \\ 0 & k_\omega \end{bmatrix} + \begin{bmatrix} \mathrm{sign}(S_\phi) \\ \mathrm{sign}(S_\omega) \end{bmatrix}, \tag{24}$$

with $k_\phi > 0$ and $k_\omega > 0$.

Finally, under the action of the control voltages the tracking errors of rotor flux and speed converges exponentially towards zero.

In the next section, the design of two controllers based on fuzzy logic is detailed to replace the two conventional equivalent commands and thus to achieve faster and more robust performance during the convergence phase.

3.3 EQUIVALENT FUZZY LOGIC CONTROL

In the fuzzy sliding mode approach proposed, the equivalent fuzzy control structures, combined with the switching control, are designed according to the fuzzy type of Mamdani.

Fuzzy logic imitates human reasoning by exploiting the experience of an expert, or the intuition of a designer. This knowledge is expressed by means of several rules of inference manipulating fuzzy operators, applied to linguistic variables [17]. The membership functions of the associated input and output variables are predefined on a universe of the discourse, and contrary to the classical logic, they can take any value between 0 and 1.

The scheme of this fuzzy controller class uses three functional blocks: a fuzzification interface, a knowledge base and a defuzzification interface [8–17]. Fuzzification consists to quantifying the input and output variables crisp into fuzzy variables by assigning to each of them fuzzy subsets designated by linguistic labels. Fuzzy subsets are represented by membership functions with predefined shapes on a universe of discourse. The knowledge base contains a collection of inference rules, in the (If-Then) format, characterizing the relationships between the input and output variables. The chosen inference method allows calculating the resulting fuzzy subset. The defuzzification interface transforms the resultant fuzzy subset into a net value and thus deduces the precise output corresponding to the appropriate control action [8].

For the FSMC controller design, two fuzzy-PI with identical configurations are used to achieve the equivalent commands v_{sdeq} and v_{sqeq} as illustrated by the Fig. 2.

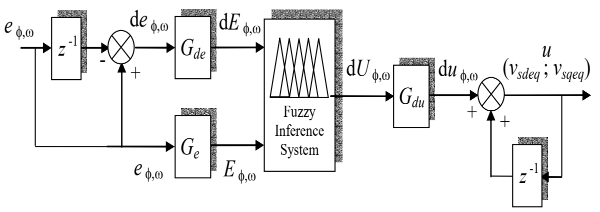

Fig.2 – Structure of fuzzy logic controller.

The input variables of the two controllers are the errors (e_ω,e_ϕ) and the variations of these errors (de_ω,de_ϕ), for both rotor flux control and speed.

Inputs and outputs are normalized in the range [–1,1] by introducing a set of scaling factors ($G_{e\phi}$,$G_{de\phi}$), ($G_{e\omega}$,$G_{de\omega}$), and ($G_{du\phi}$,$G_{du\omega}$). So, at discrete time k, normalized inputs are determined as:

$$\begin{cases} E_\phi(k) = G_{e\phi}\, e_\phi(k) \\ \mathrm{d}E_\phi(k) = G_{de\phi}\,[e_\omega(k) - e_\phi(k-1)] \end{cases}. \tag{25}$$

$$\begin{cases} E_\omega(k) = G_{e\omega}\, e_\omega(k) \\ \mathrm{d}E_\omega(k) = G_{de\omega}\,[e_\omega(k) - e_\omega(k-1)] \end{cases}. \tag{26}$$

Values of the equivalent controls v_{sdeq} (k) and v_{sqeq} (k) at the time k are obtained by adding the increments ($G_{du\phi}$.du_ϕ) and ($G_{du\omega}$.du_ω) to their previous values at (k–1).

$$\begin{cases} v_{sd\,eq}(k) = v_{sd\,eq}(k-1) + G_{\mathrm{d}u\phi}\, \mathrm{d}U_\phi(k) \\ v_{sq\,eq}(k) = v_{sd\,eq}(k-1) + G_{\mathrm{d}u\omega}\, \mathrm{d}U_\omega(k) \end{cases}. \tag{27}$$

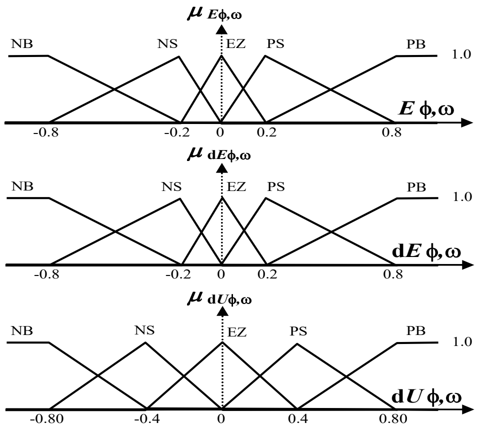

Fig.3 – Membership functions of errors, change of errors and outputs increments for both fuzzy controllers.

After several tests, identical choices were adopted for the two fuzzy regulators. Inputs and outputs are thus fuzzified into 5 fuzzy subsets, namely: Positive Big (PB), Positive Small (PS), Equal Zero (EZ), Negative Small (NS), and Negative Big (NB). The membership functions are chosen triangular and trapezoidal. This implies defining 25 rules formulated in Table 1 and expressed for example as:

$$\text{If } e(k) \text{ is NB And } \mathrm{d}e(k) \text{ is PB, Then } \mathrm{d}u(k) \text{ is PS}. \tag{28}$$

Table I

Control rules for proposed system

d*u*		d*e*				
		NB	**NS**	**EZ**	**PS**	**PB**
e	**NB**	NB	NB	NB	NS	EZ
	NS	NB	NB	NS	EZ	PS
	EZ	NB	NS	EZ	PS	PB
	PS	NS	EZ	PS	PB	PB
	PB	EZ	PS	PB	PB	PB

The reasoning method "Max-Min" of Mamdani is used for the numerical treatment of inferences and the defuzzification of the fuzzy output is based on the center of gravity technique [17]. Finally, the Fig. 4 illustrates the block diagram for the induction motor of the proposed control based the fuzzy sliding mode approach.

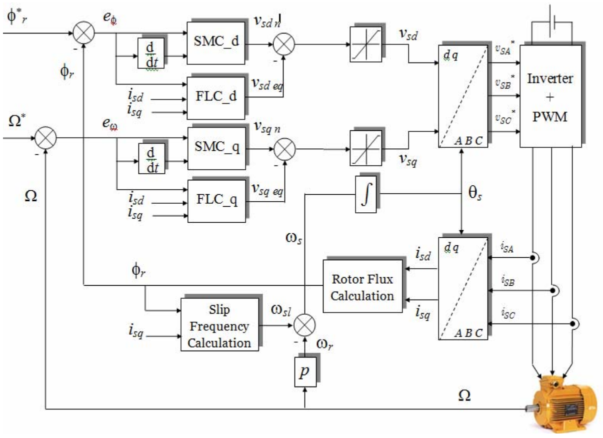

Fig.4 – Control scheme based on the FSMC approach proposed.

4. RESULTS AND DISCUSSION

In order to verify the validity and effectiveness of fuzzy sliding mode method proposed for the induction motor control, some simulation tests have been performed. The obtained performances were also compared with those of the conventional sliding mode control.

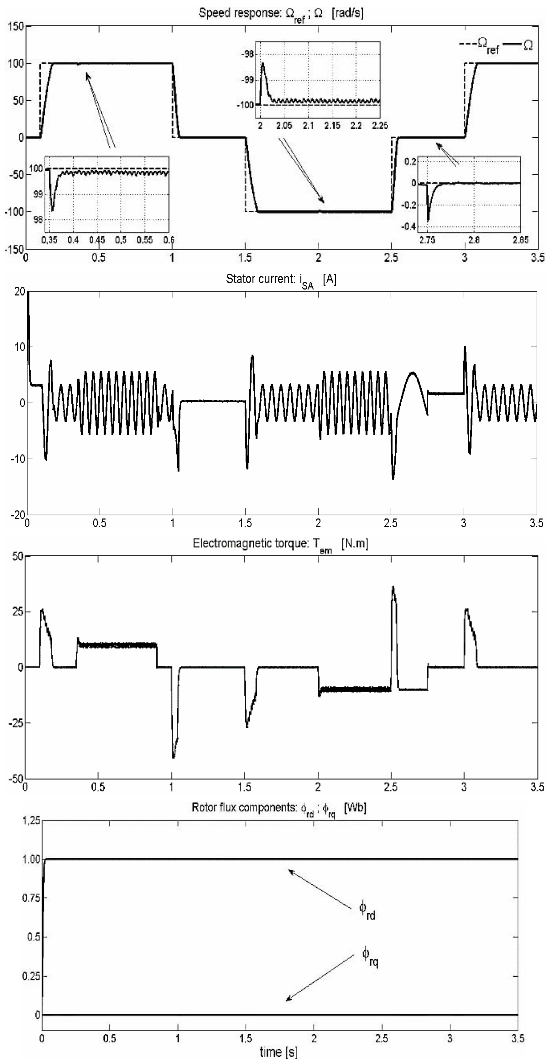

Fig.5 – Performance of conventional SMC approach: Tracking trajectory and disturbance rejection performances.

The rated values and parameters of the IM are follows: 1.5 kW, 220/380 V, 6.5/3.75 A, p = 2, 50 Hz, 1420 rpm, R_s = 4.85 Ω, R_r = 3.085 Ω, L_s = 0.274 H, L_r = 0.274 H, L_m = 0.258 H, J = 0.031 kg.m^2, B = 0.00114 kg.m/s, ϕ_r = 1 Wb.

The parameters of the SMC are: k_ϕ = 600, k_ω = 300, and those of the FSMC: k_ϕ = 600, k_ω = 300, $G_{e\phi}$ = 1/250, $G_{de\phi}$ = 1/10, $G_{du\phi}$ = 20; $G_{e\omega}$ = 1/300, $G_{de\omega}$ = 1/30, $G_{du\omega}$ = 20.

The first test has for object the study of controllers behaviours in trajectory tracking and their ability to reject disturbances. The reference benchmark defines present a fluxing phase, then a variable speed is imposed with rotation reversals between +100 and –100 rad/s and zero speed operating periods. The nominal load disturbance torque is suddenly applied and removed at different times.

Figures 5 and 6 depict respectively results obtained for conventional SMC and FSMC approaches. A comparison of speed responses is shown in Fig. 7. It can be seen that the dynamics of the FSMC has faster tracking characteristics than the conventional controller, without overshoot and better rejection of load disturbances. Oscillations due to the chattering are significantly reduced and the steady-state error is clearly lower.

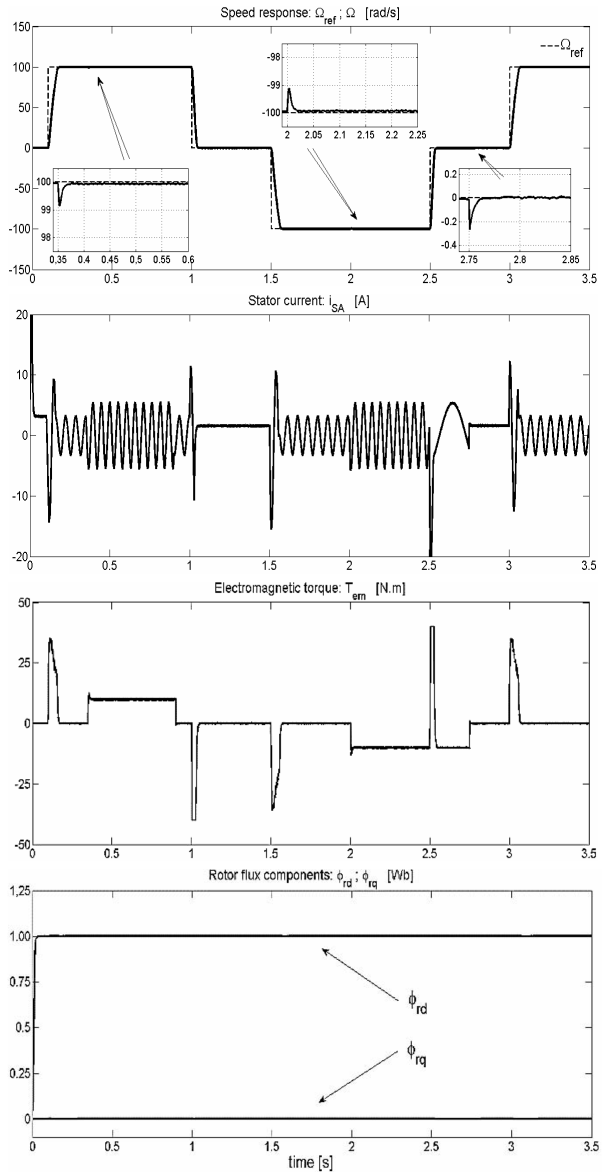

Fig.6 – Performance of hybrid FSMC approach: Tracking trajectory and disturbance rejection performances.

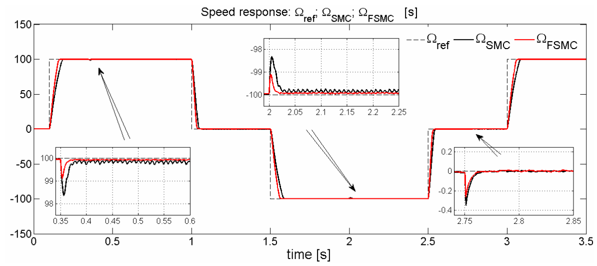

Fig.7 – Comparison of speed responses: SMC versus FSMC.

The second test concerns impact of the rotor resistance variations to the both techniques. Indeed, it is known that this resistance is the parameter that most deteriorates the robustness of the induction machine control [1]. The Figures 8 and 9 show the behavior of the two controls, when R_r is 10 % and 20 % increased of its nominal value, during a no-load starting with a 100 rad/s reference speed and a nominal load torque applied between 0.35 s and 0.9 s.

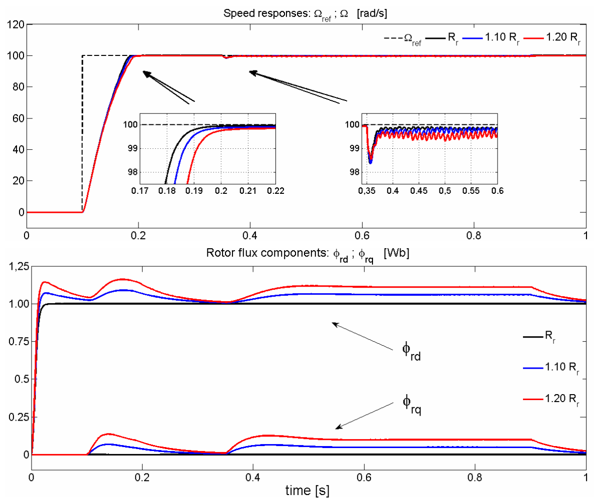

Fig.8 – Conventional SMC approach: Speed and fluxes responses for 10 % and 20 % increases in the nominal value of the rotor resistor.

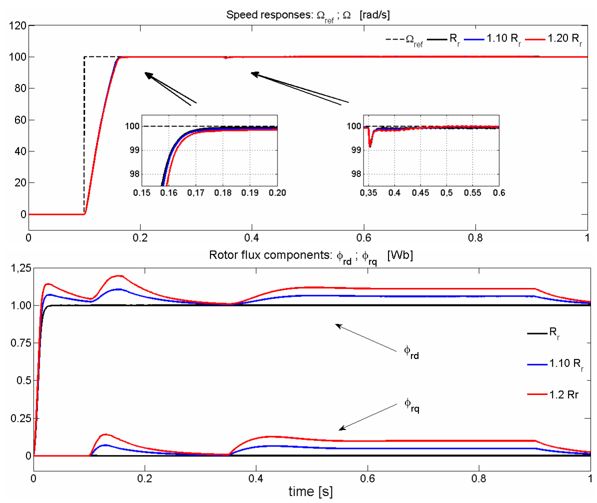

Fig.9 – Hybrid FSMC approach: Speed and fluxes responses for 10 % and 20 % increases in the nominal value of the rotor resistors.

The performances of the FSMC are practically identical when the resistance is modified, while that of the SMC changes. Thus, the robustness of the electric drive is improved by the combination of fuzzy logic and sliding modes. However, in the transient states, orientation of rotor flux is altered for both techniques.

5. CONCLUSION

In this paper, a non-linear control for an induction motor has been presented. The proposed control law is based on a combination of the sliding mode technique and the fuzzy logic approach. The improvements can be observed for trajectory tracking in the steady and transient states, robustness against changes in rotor resistance and load disturbances is better. But the major advantage of this method is the reduction of HF oscillations caused by the chattering. Simulation results show the feasibility of the method and highlight the interest of the proposed technique for different operating conditions.

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