American Journal of Applied Sciences 4(12): 987-994, 2007
ISSN 1546-9239

A Sliding Mode Speed Control of an Induction Motor

Kais JAMOUSSI, Mohamed OUALI and Hassen CHARRADI
Control Motor and Power Network (CMERP)
Engineering National School, University of Sfax (ENIS), BP W, 3038 Sfax, Tunisia

Abstract: Field-oriented control was combined to robust sliding mode for motor speed control. A smooth continuous function was added in order to overcome chattering caused by Sliding Mode Controller (SMC). Simulation results showed that improvement made by our approach compared to classical PID control.

Keywords Sliding mode control, Field-oriented control, Induction motor

INTRODUCTION

Induction motors were widely used in industry due to their relatively low cost and high reliability. To get high performances of these motors, several control approaches had been developed [1, 2, 3, 4, 5] and they can be divided in two classes: Field Oriented Control (FOC) and Direct Torque Control (DTC). The principle of DTC for induction motor drives consist of controlling simultaneously stator flux and electromagnetic torque of the machine [7, 8]. The DTC produces fast response and robust control AC adjustable speed drives [9]. As for with FOC technique, the rotor speed was asymptotically decoupled from rotor flux like DC motor and the speed depends linearly of torque current [10, 19]. However, the performance will be degraded face to motor parameter variations or unknown external disturbances [11]. To offer control robustness with minimum complexity many strategies have been proposed in literature. See for example, the adaptive control in [3, 12], fuzzy control in [1, 2, 4], robust control in [13] and sliding mode control (SMC) in [14, 15, 16]. Recently, intelligent control had been proposed and widely applied to many control application. See for example the neural control for AC motors [17]. A combination of fuzzy control and SMC, called Fuzzy Sliding mode Control, had been proposed to overcame some problems see for example[1, 2, 18]. Even so, obtaining good performances by the classical FOC of induction motors was not evident due to difficulties encountered in the determination of motor parameters and the construction of the observers. In order to overcome these difficulties, the Sliding Mode Control (SMC) was adopted [23, 14, 20, 16], because it has many good features, such as robustness to parameter variations or load disturbances, fast dynamic response, and simplicity of design and implementation.

During sliding mode functioning, the system was forced to slide along or near the vicinity of the switching surface [21]. The system becomes then robust and insensitive to the interaction, disturbance and variation of parameters. Moreover, the SMC mode does not require an accurate model of the motor; this might necessitate only information on parameters value.

Yet, the insensitivity of the controlled system against uncertainties was guaranteed during the steady operation [12]. So, the system robustness cannot be maintained in the whole control process. To solve this problem, a preparation of the induction machine to the field–oriented control was suggested. The Proposed method includes two phases. The first one concerns the flux establishment in size and in direction, while the second was used to maintain the flux already settled by regulating the direct component of stator current. Since, the flux was constant inside the machine, a speed consign was applied. Then, speed control ensured by the sliding mode.

Moreover, it was well known that sliding mode techniques generate undesirable chattering. This problem can be remedied by replacing the switching function by a smooth continuous function [22, 27].

This paper proposes a new control FOC based algorithm using SMC for an induction motor drive.

In a second section, a description of the machine was given. The third section treats the preparation of the motor to the FOC. Sections four and five present the direct component regulation of stator current and the angular speed by means of sliding mode control.

Finally, in section six the simulation results were presented and discussed.

MATERIALS AND METHODS

Description of the induction motor: Before the implementation of any control mode, it was necessary to define the function equations.

For this purpose, an induction motor model was established using a rotating field reference *(d,q)* (without saturation).

The equation system of the stator was [23][24][6]:

$$\begin{cases} V_{ds} = R_s I_{ds} + \frac{d}{dt}\psi_{ds} - \omega_s \psi_{qs} \\ V_{qs} = R_s I_{qs} + \frac{d}{dt}\psi_{qs} + \omega_s \psi_{ds} \end{cases} \quad (1)$$

As for the stator, the equation system was :

$$\begin{cases} 0 = R_s I_{dr} + \frac{d}{dt}\psi_{dr} - \omega_r \psi_{dr} \\ 0 = R_s I_{qr} + \frac{d}{dt}\psi_{qr} + \omega_r \psi_{qr} \end{cases} \quad (2)$$

Where V_{sd} and V_{sq} were the stator voltage components, I_{sd} and I_{sq} were the stator current components, ψ_{qs} and ψ_{ds} were the stator flux components, ψ_{qr} and ψ_{dr} were the rotor flux components. R_s and R_r were respectively the stator and rotor resistances.

Finally the mechanical equation was given by :

$$J\frac{d}{dt}\omega_m = T_e - T_r \quad (3)$$

where *J* represent the moment of inertia.

Hence, the electromagnetic torque equation can be witten in terms of ψ_{dr} and ψ_{qr} as :

$$T_e = \frac{L_m}{L_r + L_m}(\psi_{dr} I_{qs} - \psi_{qr} I_{ds}) \quad (4)$$

Where $\omega_r = \omega_s - \omega_m$

The stator and rotor flux linkages expressions were [25]:

$$\begin{cases} \psi_{ds} = L_s I_{ds} + L_m I_{dr} \\ \psi_{qs} = L_s I_{qs} + L_m I_{qr} \\ \psi_{dr} = L_r I_{dr} + L_m I_{ds} \\ \psi_{qr} = L_r I_{qr} + L_m I_{qs} \end{cases} \quad (5)$$

Where L_s and L_r were stator and rotor cyclic inductances, L_m was the mutual inductance.

The rotor currents were given by: [23][25]

$$\begin{cases} I_{dr} = \frac{1}{L_r}\psi_{dr} - \alpha I_{ds} \\ I_{qr} = \frac{1}{L_r}\psi_{qr} - \alpha I_{qs} \end{cases} \quad (6)$$

Where $\alpha = \frac{L_m}{L_r}$

While writing Eqs (6) in (2) and assuming $\tau_r = L_r / R_r$, the rotor flux equations were expressed as :

$$\begin{cases} \dot{\psi}_{dr} = -\frac{1}{\tau_r}\psi_{dr} + \omega_r \psi_{qr} + \frac{L_m}{\tau_r} I_{ds} \\ \dot{\psi}_{qr} = -\omega_r \psi_{dr} - \frac{1}{\tau_r}\psi_{qr} + \frac{L_m}{\tau_r} I_{qs} \end{cases} \quad (7)$$

Using Eqs (8) and Eqs (5), the stator flux was given by:

$$\begin{cases} \psi_{ds} = (L_s - \alpha L_m) I_{ds} + \alpha \psi_{dr} \\ \psi_{qs} = (L_s - \alpha L_m) I_{qs} + \alpha \psi_{qr} \end{cases} \quad (8)$$

Some notations were considered to get some simplification, so, we consider:

$L_1 = L_s - \alpha L_m$ and $R_1 = R_s + \alpha^2 R_r$.

Differentiating Eqs. (8), substituting them into Eqs. (1) and letting $\tau_1 = L_1 / R_1$ leads to :

$$\begin{cases} \dot{I}_{ds} = \frac{\alpha}{\tau_1 L_1}\psi_{dr} + \frac{\alpha}{L_1}\omega_m \psi_{qr} - \frac{1}{\tau_1} I_{ds} + \omega_s I_{qs} + \frac{1}{L_1} V_{ds} \\ \dot{I}_{qs} = -\frac{\alpha}{L_1}\omega_m \psi_{dr} + \frac{\alpha}{\tau_1 L_1}\psi_{qr} - \omega_s I_{ds} + \frac{1}{\tau_1} I_{qs} + \frac{1}{L_1} V_{qs} \end{cases} \quad (9)$$

Finally we get the state equations of the induction machine using. (7), (9) and (3) :

$$\begin{cases} \dot{\psi}_{dr} = -\frac{1}{\tau_r}\psi_{dr} + \omega_r \psi_{qr} + \frac{L_m}{\tau_r} I_{ds} \\ \dot{\psi}_{qr} = -\omega_r \psi_{dr} - \frac{1}{\tau_r}\psi_{qr} + \frac{L_m}{\tau_r} I_{qs} \\ \dot{I}_{ds} = \frac{\alpha}{\tau_1 L_1}\psi_{dr} + \frac{\alpha}{L_1}\omega_m \psi_{qr} - \frac{1}{\tau_1} I_{ds} + \omega_s I_{qs} + \frac{1}{L_1} V_{ds} \\ \dot{I}_{qs} = -\frac{\alpha}{L_1}\omega_m \psi_{dr} + \frac{\alpha}{\tau_1 L_1}\psi_{qr} - \omega_s I_{ds} + \frac{1}{\tau_1} I_{qs} + \frac{1}{L_1} V_{qs} \\ \dot{\omega}_m = \frac{1}{J}[T_e - T_m] \end{cases} \quad (10)$$

Since it was fully described by the equation system (10), the motor was ready to be controlled using FOC method. Nevertheless, it can be seen that equation system 10 describes the motor by complicate and non linear model. So, it was necessary to adopt the decoupling relationship means of a proper selection of state coordinates, under the simplifying hypothesis that the rotor flux was kept constant [23].

Preparation of motor to the Field Oriented Control: Since, the decoupling condition of the feedback linearization method always holds, the parameters of the induction motor must be precisely known and accurate information on the rotor flux was required. Meantime, the dynamic behavior of the induction motor under the field-oriented or feedback linearization control techniques was quite similar to that of a separately excited DC motor, and the control effort was simplified. However, the performance will be degraded due to the motor parameter variations or unknown external disturbances.

To prepare the machine to the field-oriented control, two conditions should be satisfied.

- The module of the vector of flux must be constant and equal to the nominal value.
- The direction must be confused with the direct axis "d".

What was to say that the rotor flux quadrature component was equal to zero.

$$\psi_{qr} = 0 \text{ or } \dot{\psi}_{qr} = 0 \quad with \quad \psi_{qr}(0) = 0$$

This allows eliminate the terms ψ_{qr} in the state Eqs. (10), which reduces the model order to four. Indeed, the second state equation of the system (10) becomes equal to :

$$\omega_s = \omega_m + \frac{L_m I_{qs}}{\tau_r \psi_{dr}} \tag{11}$$

Which give the expression of the stator pulsation according to the variables of state gives. Then the equations of state become:

$$\begin{cases} \dot{\psi}_{dr} = -\frac{1}{\tau_r}\psi_{dr} + \frac{L_m}{\tau_r} I_{ds} \\ \dot{I}_{ds} = \frac{\alpha}{\tau_1 L_1}\psi_{dr} - \frac{1}{\tau_1} I_{ds} + \omega_s I_{qs} + \frac{1}{L_1} V_{ds} \\ \dot{I}_{qs} = -\frac{\alpha}{L_1}\omega_m \psi_{dr} - \omega_s I_{ds} + \frac{1}{\tau_1} I_{qs} + \frac{1}{L_1} V_{qs} \\ \dot{\omega}_m = \frac{1}{J}[T_e - T_m] \\ T_e = (\psi_{dr} I_{qs} - \psi_{qr} I_{ds}) \end{cases} \tag{12}$$

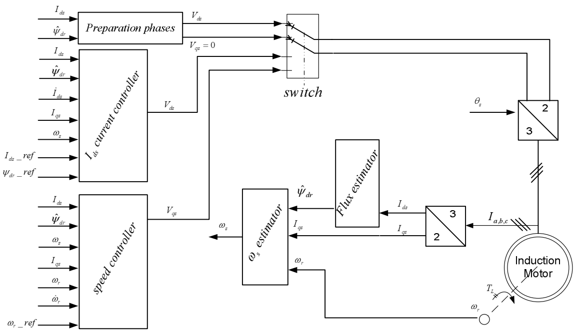

Fig.1: Bloc diagram of the field oriented control and the preparation phases

Flux establishment: In the equation system (12) it was easy to notice that the only control which acts on ψ_{dr}, while $\omega_s = 0$, was V_{ds} through the stator current I_{ds}. In order to put flux to its nominal value in a minimum time ($\psi_{dr} = \psi_{drN}$) it was necessary [26]:

- To apply the maximum value of the control ($V_{ds} = V_{ds\max}$).
- To maintain $\omega_s = 0$ to eliminate the interaction between the currents I_{ds} and I_{qs}. That was done by annulling of the effect of the terms $\omega_s I_{qs}$ and $\omega_s I_{ds}$ which appear in the second and the third equation of system (12). So we obtain two completely decoupled systems. The system of equations (12) will be reduced to its the first two equations was :

$$\begin{cases} \dot{\psi}_{dr} = -\frac{1}{\tau_r}\psi_{dr} + \frac{L_m}{\tau_r} I_{ds} \\ \dot{I}_{ds} = \frac{\alpha}{\tau_1 L_1}\psi_{dr} - \frac{1}{\tau_1} I_{ds} + \frac{1}{L_1} V_{ds} \end{cases} \tag{13}$$

Since $\tau_r << \tau_1$ (section 2), the application of the control ($V_{ds} = V_{ds\max}$) generates a fast variation of I_{ds} current. This later reaches its nominal value long before ψ_{dr}. Thus, the application of such command must be organized as follows:

a) As the machine coil accepts a current up to 1.2 of the nominal value I_{dsN}, it becomes possible to apply the control ($V_{ds} = V_{ds\max}$) until I_{ds} arrives to 1.2 I_{dsN}, which permits to accelerate the phase of flux establishment.

b) To maintain I_{ds} to 1.2*I_{dsN} which involves $\dot{I}_{ds} = 0$ and so V_{ds} expression becomes:

$$V_{ds} = -\frac{\alpha}{\tau_r}\psi_{dr} + \frac{L_1}{R_1} I_{ds}. \tag{14}$$

Consequently, the first Eq. of the system (13) will be reduced to a first order equation (15):

$$\dot{\psi}_{dr} = -\frac{1}{\tau_r}\psi_{dr} + 1.2\frac{L_m}{\tau_r} I_{dsN} \tag{15}$$

Hence, the flux rises according to the time constant τ_r

c) When the nominal value of ψ_{dr} was reached, the following control V_{ds} was applied:

$$V_{ds} = -\frac{\alpha}{\tau_1}\psi_{drN} + R_1 I_{dsN} \tag{16}$$

That permits to maintain ψ_{dr} at its nominal value. Thus $\dot{\psi}_{dr} = 0$ and using eq. (13), $\psi_{drN} = L_m I_{dsN}$. Eq. (16) can be written:

$$V_{ds} = (R_1 - \alpha^2 R_r) I_{dsN}. \tag{17}$$

This phase was stopped when I_{ds} decreases to I_{ds_ref}.

In order to regulate the flux ψ_{dr} around its reference value (0.85Wb), the I_{ds} current was maintained to its reference value.
The machine state equations were given by:

$$\begin{cases} \dot{I}_{qs} = -\dfrac{\alpha}{L_1}\omega_m \psi_{dr} - \omega_s I_{ds} + \dfrac{1}{\tau_1} I_{qs} + \dfrac{1}{L_1} V_{qs} \\ \dot{I}_{ds} = \dfrac{\alpha}{\tau_1 L_1}\psi_{dr} - \dfrac{1}{\tau_1} I_{ds} + \omega_s I_{qs} + \dfrac{1}{L_1} V_{ds} \\ \dot{\psi}_{dr} = -\dfrac{1}{\tau_r}\psi_{dr} + \dfrac{L_m}{\tau_r} I_{ds} \\ \dot{\omega}_m = \dfrac{1}{J}\left[\psi_{dr} I_{qs} - T_m\right] \end{cases} \quad (18)$$

Regulation of Ids current: Equation 18.b, gives the expression of ψ_{dr} as:

$$\psi_{dr} = \frac{\tau_1 L_1}{\alpha}(\dot{I}_{ds} + \frac{1}{\tau_1} I_{ds} - V) \quad (19)$$

where $V = \omega_s I_{qs} + \frac{1}{L_1} V_{ds}$ was the control input function.

Equation (19) shows that ψ_{dr} depends only on the I_{ds} current.
The derivative of Eq. 18.b was:

$$\ddot{I}_{ds} = \frac{\alpha}{\tau_1 L_1}\dot{\psi}_{dr} - \frac{1}{\tau_1}\dot{I}_{ds} + \dot{V}_1 \quad (20)$$

By substituting Eq. 18.c and Eq.(19) in Eq. (20), we obtain

$$\begin{aligned} \ddot{I}_{ds} &= -(\frac{1}{\tau_r} + \frac{1}{\tau_1})\dot{I}_{ds} - (\frac{1}{\tau_r \tau_1} - \frac{L_m \alpha}{\tau_r^{\,2} L_1}) I_{ds} + U_f \\ &= -A_f \dot{I}_{ds} - B_f I_{ds} + U_f \end{aligned} \quad (21)$$

Where $U_f = \frac{1}{\tau_r} V + \dot{V}$ was the new control function.

Let $e_i = I_{ds} - I_{ds\_ref}$: the error on I_{ds} current. We can write: $\dot{e}_i = \dot{I}_{ds}$.

Application of the sliding mode: The implementation of a variable structure regulator consists on the application of a control function (describe in the system (22)) for render the system insensitive against parameters variations [23, 20].

$$U_i = \begin{cases} U_i^{+}(x,t) & if \quad S_i > 0 \\ U_i^{-}(x,t) & if \quad S_i < 0 \end{cases} \quad (22)$$

Where S_i was the ith component of a switching surface $S_i = 0$.
This system with discontinuous control was called variable structure system, since the control structure switches alternatively according to the state of the system. The sliding mode occurs on a switching surface $S_i(x) = 0$, which forces the original system to behave as linear time invariant system, which can be considered to be stable.
In our study the surfaces were taken to be linear and written as :

$$S_f = c_1 e_i + \dot{e}_i \quad (23)$$

$c_1 = constant$

The design of such a system involves [26][23]:

(i) The choice of the control function U_i to guarantee the existence of a sliding mode.
(ii) The determination of the switching function S such that the system has the desired response.
(iii) The elimination of chattering of the control signal.

Control function

$\dot{S}_f$ was given by

$$\dot{S}_f = -(A_f - c_1)\dot{I}_{ds} - B_f I_{ds} + U_f \quad (24)$$

Let

$$\begin{cases} A_f = A_{fn} + \Delta A_f \\ B_f = B_{fn} + \Delta B_f \end{cases}$$

Here A_{fn} and B_{fn} were the nominal values, and ΔA_f and ΔB_f were the associated variations.
Let the control function U_f be decomposed into

$$U_f = U_{f\,eq} + \Delta U_f \quad (25)$$

Where $U_{f\,eq}$, called the equivalent control, was defined as the solution of the problem $S' = 0$ under $A_f = A_{fn}$, $B_f = B_{fn}$. That was :

$$U_{f\,eq} = (A_{fn} - c_1)\dot{I}_{ds} + B_{fn} I_{ds} \quad (26)$$

The function ΔU_f was the discontinuous term, employed to eliminate the influence due to ΔA_f and ΔB_f so as to guarantee the existence of sliding mode was constructed as :

$$\Delta U_f = \lambda_{f1}\dot{I}_{ds} + \lambda_{f2} I_{ds} + \lambda_{f3}\left|S_f\right| \quad (27)$$

It was known that the condition of the existence and reachability of a sliding motion was [22, 27]

$$\dot{S}_f S_f < 0 \quad (28)$$

Substitution of Eqs (25),(26) and (27) into Eqn (24) yields

$$\dot{S}_f = (\lambda_{f1} - \Delta A_f)\dot{I}_{ds} + (\lambda_{f2} - \Delta B_f) I_{ds} + \lambda_{f3}\left|S_f\right| \quad (29)$$

Then

$$\begin{aligned} S_f \dot{S}_f &= (\lambda_{f1} - \Delta A_f)\dot{I}_{ds} S_f + (\lambda_{f2} - \Delta B_f) I_{ds} S_f \\ &\quad + \lambda_{f3}\left|S_f\right| S_f \end{aligned} \quad (30)$$

Then the condition for satisfying inequality $\dot{S}_f S_f < 0$ were :

$$\lambda_{f1} = \begin{cases} \lambda^+_{f1} < \min(\Delta A_f) & if \quad S_f \dot{I}_{ds} > 0 \\ \lambda^-_{f1} > \max(\Delta A_f) & if \quad S_f \dot{I}_{ds} < 0 \end{cases}$$

$$\lambda_{f2} = \begin{cases} \lambda^+_{f2} < \min(\Delta B_f) & if \quad S_f I_{ds} > 0 \\ \lambda^-_{f2} > \max(\Delta B_f) & if \quad S_f I_{ds} < 0 \end{cases} \qquad (31)$$

$$\lambda_{f3} = \begin{cases} \lambda^+_{f3} & if \quad S_f > 0 \\ \lambda^-_{f3} & if \quad S_f < 0 \end{cases}$$

Then the quantities λ_{f1}, λ_{f2} and λ_{f3} could be chosen as follows

$$\begin{cases} \lambda_{f1} = \lambda^+_{f1} = -\lambda^-_{f1} \\ \lambda_{f2} = \lambda^+_{f2} = -\lambda^-_{f2} \\ \lambda_{f3} = \lambda^+_{f3} = -\lambda^-_{f3} \end{cases} \qquad (32)$$

The control law ΔU_f can be expressed as

$$\Delta U_f = (\lambda_{f1}\dot{I}_{ds} + \lambda_{f2} I_{ds} + \lambda_{f3}|S_f|)\operatorname{sgn}(S_f) \qquad (33)$$

Regulation of angular speed: The equations used for regulating the angular speed were written in the following

$$\dot{I}_{qs} = -\frac{\alpha}{L_1}\omega_m \psi_{dr} - \omega_s I_{ds} + \frac{1}{\tau_1} I_{qs} + \frac{1}{L_1} V_{qs} \qquad (34)$$

$$\dot{\omega}_m = \frac{1}{J}\left[\frac{L_m}{L_m + L_r}(\psi_{dr} I_{qs}) - T_m\right] \qquad (35)$$

From the Eqs (35) we wirte :

$$\ddot{\omega}_m = \frac{1}{J}\left[\frac{L_m}{L_m + L_r}(\psi_{dr}\dot{I}_{qs}) - \dot{T}_m\right] \qquad (36)$$

The elimination of the state variable I_{qs} and by the substitution of Eqs (35), (34) in Eqs (36) gives the following differential equation in ω_m

$$\ddot{\omega}_m = -\frac{1}{\tau_1}\dot{\omega}_m - \frac{L_m \alpha \psi^2_{dr}}{J(L_m + L_r)L_1}\omega_m - \frac{1}{J\tau_1}T_m + \frac{L_m \psi_{dr}}{J(L_m + L_r)}\left(\frac{V_{qs}}{L_1} - \omega_s I_{ds}\right) \qquad (37)$$

Or $\ddot{\omega}_m = -A_v \dot{\omega}_m - B_v \omega_m - C_v T_m + U_v$ (38)

Where $U_v = \frac{L_m \psi_{dr}}{J(L_m + L_r)}\left(\frac{V_{qs}}{L_1} - \omega_s I_{ds}\right)$ represent the control input.

The error on the angular speed was written $e_v = \omega_m - \omega_{ref}$ and $\dot{e}_v = \dot{\omega}_m$

Where ω_{ref} was the desired angular speed.

In the same way, the switching surface was chosen:

$$S_v = c_2 e_v + \dot{e}_v \qquad (39)$$

Let

$$\dot{S}_v = c_2 \dot{e}_v + \ddot{e}_v = c_2(\omega_m - \omega_{ref}) + \ddot{\omega}_m \qquad (40)$$

And we substitute $\ddot{\omega}_m$ by it expression, we obtain :

$$\dot{S}_v = -(A_v - c_2)\dot{\omega}_m - B_v \omega_m - C_v T_m + U_v \qquad (41)$$

Let us consider

$$\begin{cases} A_v = A_{vn} + \Delta A_v \\ B_v = B_{vn} + \Delta B_v \\ C_v = C_{vn} + \Delta C_v \end{cases}$$

Let the control function U_v was constructed as

$$U_v = U_{veq} + \Delta U_v \qquad (42)$$

The equivalent control $U_{v\,eq}$ was build as follows

$$U_{veq} = (A_v - c_2)\dot{\omega}_m + B_v \omega_m + C_v T_m \qquad (43)$$

And the discontinuous control was

$$\Delta U_v = \lambda_{v1}\dot{\omega}_m + \lambda_{v2}\omega_m + \lambda_{v3}|S_v| \qquad (44)$$

Substituting Eqn. (42), (43) and (44) in (41), we obtain

$$\dot{S}_v = (\lambda_{v1} - \Delta A_v)\dot{\omega}_m + (\lambda_{v2} - \Delta B_v)\omega_m + \lambda_{v3}|S_v| \qquad (45)$$

To satisfy the sliding condition $\dot{S}_v S_v < 0$ [23], it was sufficient to take :

$$\lambda_{v1} = \begin{cases} \lambda^+_{v1} < \min(\Delta A_v) & if \quad S_v \dot{\omega}_m > 0 \\ \lambda^-_{v1} < \max(\Delta A_v) & if \quad S_v \dot{\omega}_m < 0 \end{cases}$$

$$\lambda_{v2} = \begin{cases} \lambda^+_{v2} < \min(\Delta B_v) & if \quad S_v \omega_m > 0 \\ \lambda^-_{v2} < \max(\Delta B_v) & if \quad S_v \omega_m < 0 \end{cases} \qquad (46)$$

$$\lambda_{v3} = \begin{cases} \lambda^+_{v3} < \min(\Delta C_v T_m) & if \quad S_v > 0 \\ \lambda^-_{v3} < \max(\Delta C_v T_m) & if \quad S_v < 0 \end{cases}$$

Chattering

For the control law given by Eqs. (42), if λ_{vk}, $k=1,2,3$ were chosen as : $\lambda_{vk} = \lambda^+_{vk} = -\lambda^-_{vk}$ then the discontinuous control function U_v can be represented as

$$\Delta U_v = (\lambda_{v1}\dot{\omega}_m + \lambda_{v2}\omega_m + \lambda_{v3}|S_v|)sign(S_v) \qquad (47)$$

Since the control U_v contains the sign function, direct application of such a control signal to the motor may give rise to chattering. To reduce the chattering, the sign function *sign(S_v)* in the Eqn.(47) can be replaced by modified proper continuous function [27] as :

$\frac{S_v}{|S_v| + \delta}$ where δ was chosen as a function of $|\omega_m - \omega_{ref}|$ as $\delta = \varepsilon_0 + \varepsilon_1|\omega_m - \omega_{ref}|$ and the value of ε_0 and ε_1 were positive constants.

RESULTS AND DISCUSSION

The Simulink/Matlab software package implementation was adopted because its inherent integration of vectorized system representations, time graphical portrayal of signals combined evolutions and the simple functionality realization of the controllers and power electronic excitations.
The motor parameters were given in Table 1 and they were identified in the last work [28].

Table 1: Induction motor parameters

Parameters	Value
R_r	29Ω
R_s	10 Ω
L_r	680mH
L_s	690mH
L_m	660mH
J	0.00062 Kg.m²
Number of Poles P	1 Pair of poles
Power	370W

The simulation of the inputs to the machines involves the mathematical representation of programmed time sequence of events such as the sudden application or removal of mechanical loads, the ramping of the magnitude and frequency of the applied voltages, or even the changes in parameter values (for instance R_r and R_s).

Preparation phases : Fig. 3 and 4 show the evolution of flux and I_{ds} current during the preparation of the motor to the field-oriented control. The analysis of these figures leads to the classification of the time space into three phases. It was to be noted that during these phases, ω_s was maintained equal to zero.

1st phase : corresponds to the application of the control $V_{ds} = V_{ds\max}$ until Ids reaches 1.2 times its nominal value. In this case, the motor was described by Eqs. (13).

2nd phase : in this phase, the control law giving the expression of V_{ds} was described by Eqs (14), which corresponds to the maintaining of I_{ds} current equal to 1.2 times of its nominal value, until ψ_{dr} reaches its nominal value.

3rd phase : start at 0.12s. At this time we apply the control law leads to maintain the flux ψ_{dr} equal to its reference value. (Fig 2 to 5)

Start up: We note that we can apply the desired input speed after 0.2s (when the motor was ready).
One chose to apply the reference speed after 1s.
In order to test the robustness of the proposed method, we simulated the control strategy using PID and sliding mode. For both cases, we excited the simulator with the same input variables (consign angular speed and perturbation). Results were illustrated by Fig. 6.
It can be see that the sliding mode offers a robust control compared to PID control manly when applying a load torque to the system (from instant 2 and 4.5s)

To show the performance of the proposed structure at varying operation ranges, we introduce a random variation on the stator and rotor resistances (this internal disturbance was due to the internal temperature variation of the motor). Moreover we applies the torque load between t = 2s and 4,5s with 10% of incertitude Fig 7.
The influence of the load variation applied between t = 2s and 4.5s was illustrated on Fig.8 and 9.

It was clear that the chattering on the control signal was decreased when we replaced the sign function by a continuous function Fig. 10 and 11.

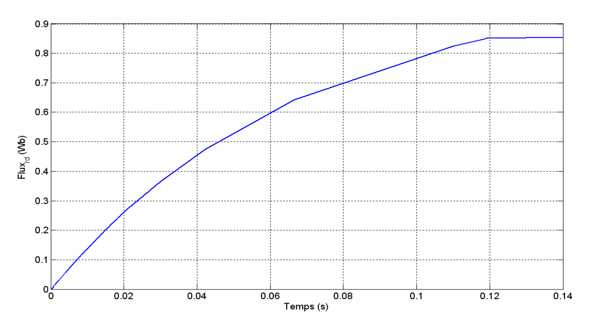

Fig. 2: Zoom of the rotor flux (the preparation phases)

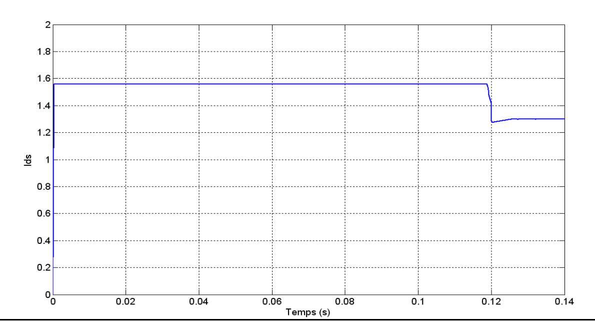

Fig. 3: Zoom of the I_{ds} current (the preparation phases)

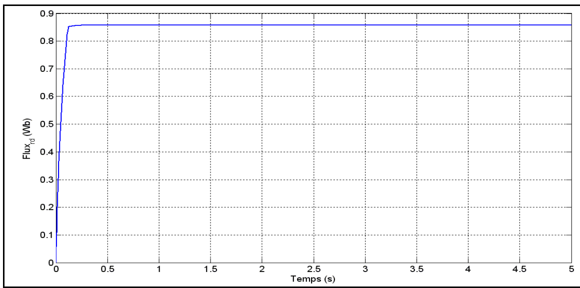

Fig. 4: Behaviour of ψ_{dr}

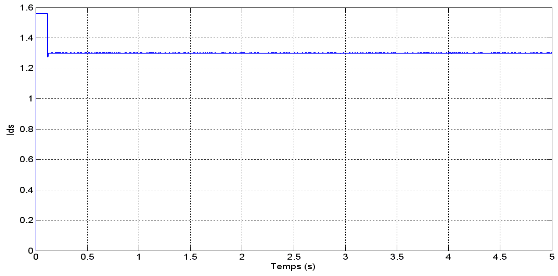

Fig. 5: Behaviour of I_{ds}

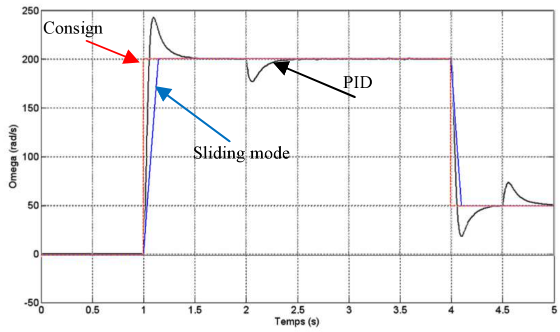

Fig. 6: Rotor speed response with sliding mode control/PID control

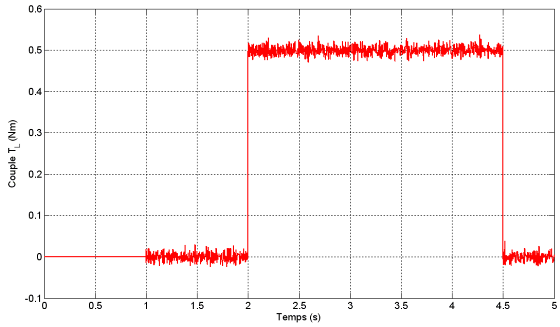

Fig. 7: Applied torque T_m

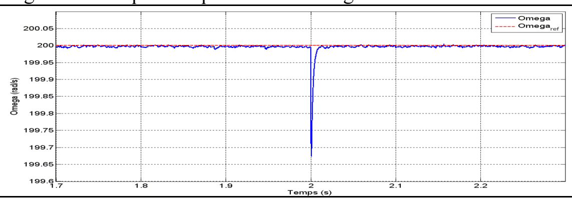

Fig. 8: Zoom of the rotor speed response when T_m was applied at t = 2s

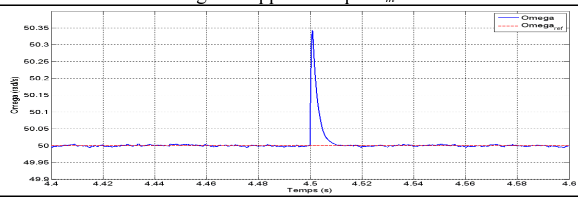

Fig. 9: Zoom of the rotor speed response when Tm was disabled at t = 4.5s

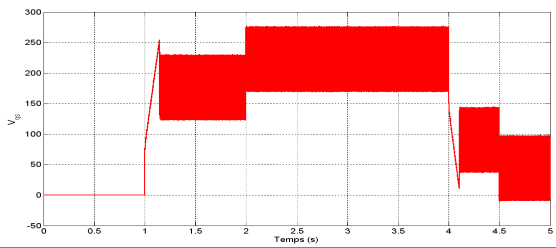

Fig. 10: Behaviour of Vds control

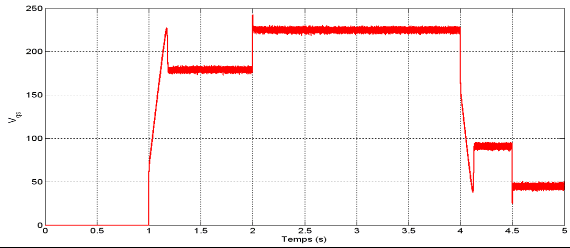

Fig. 11: Behaviour of Vds control after reducing chattering

CONCLUSION

The field-oriented control using a sliding mode speed was employed to obtain the better performance from the induction motor in a speed control. Also, compared to the classical PID control, a sliding mode approach gives a rather accurate response in face of large plant parameter variations and external disturbances.

In ours future study, we will interest to test our algorithm on an experimental plant composed of a 0.37kW induction motor. The command was designed with a digital signal processor TMS320F240 board.

REFERENCES

1. J. Wang, S. S. Get and T. H. Lee, 2000. Adaptive Fuzzy Sliding Mode Control of a Class of Nonlinear Systems. Proceeding of the 3rd Asian Control Conference, Shanghai, pp: 599-604.

2. Ta-Tau Chen and Tzuu-Hseng S. Li, 2000. Integrated Fuzzy GA-Based Simplex Sliding-Mode Control. International Journal of Fuzzy Systems, 2 (4): 267-277.
3. S. Belkhodja B. de Fornel, 1996. Commande adaptative d'une machine asynchrone. J. Phys. III, 6 : 779-796.
4. MINH TA CAO, 1997. Commande numérique de machines asychrones par logique floue, Thesis, School of Science Québec, University LAVAL, 1997.
5. Cristian Lascu, Ion Boldea and Frede Blaabjerg, 2004. Direct Torque Control of Sensorless Induction Motor Drives: A Sliding-Mode Approach. IEEE Transactions on Industry Applications, 40 (2): 582-590.
6. H. Sarhan and R. Issa, 2006. Improving Mechanical Characteristics of Inverter-induction Motor Drive System. American Journal of Applied Sciences, 3 (8): 1961-1966.
7. T. Brahmananda Reddy, J. Amarnath and D. Subba Rayudu, 2006. Direct Torque Control of Induction Motor Based on Hybrid PWM Method for Reduced Ripple: A Sliding Mode Control Approach. ACSE Journal, 6 (4): 23-30.
8. R.Toufouti S.Meziane ,H. Benalla, 2006. Direct Torque Control for Induction Motor Using Fuzzy Logic. ACSE Journal, 6 (2): 17-24.
9. I. Takahashi and Y. Ohmori, 1989. High-performance direct torque control of induction motor. IEEE Trans. Ind. Appl, 25 (2):257–264.
10. F. Blaschke, 1972. The principle of field orientation as applied to the new transvector closed-loop control system for rotating field machine. Siemens Rev., 39, pp: 217-220.
11. M. Djemai, J. Hernandez and J.P. Barbot, 1993. Nonlinear Control with Flux Observer for a Singularly Perturbed Induction Motor. Proceeding of the 32nd Conference on Decision and Control, Texas, pp: 3391-3396.
12. R. J. Wai, 2000. Adaptive sliding-mode control for induction servo motor drive. IEE Proc. Elect. Power Application 147 (6): 533-562.
13. Rong-Jong Wai and Kuo-Min Lin, 2005. Robust Decoupled Control of Direct Field-Oriented Induction Motor Drive. IEEE Transactions on Industrial Electronics, 52 (3): 837-854.
14. V. I. Utkin, 1993. Sliding mode control design principle and application to electric drives. IEEE Trans. Ind. Electron., 40 (1): 23-36.
15. L. G. Shiau and J. L. Lin, 2001. Stability of sliding-mode current control for high performance induction motor position drives. IEE Proc. Elect. Power Applicat., 148 (1): 69-75.
16. R. J. Wai and W. K. Liu, 2001. Nonlinear decoupled control for linear induction motor servo-drive using the sliding-mode technique. IEE Proc. Contr. Theory Applicat.,148 (3): 217-231.
17. M. Kuchar, P. Brandstetter and M. Kaduch, 2004. Sensorless induction motor drive with neural network. PESC, Aachen Germany, pp: 3301-3305,
18. Qinghui Wu and Cheng Shao, 2006. Novel Hybrid Sliding-mode Controller for Direct Torque Control Induction Motor Drives. Proceedings of the American Control Conference Minneapolis, Minnesota, USA, pp: 2754-2758.
19. T. Ourth, 1993. Commande vectorielle d'un moteur asynchrone sans capteur. J. Phys. III France 3, pp : 1123-1133
20. V.1 Utkin, 1978. Sliding mode and their application in variable structure systems. Moscow.
21. A. Levant, 1993. Sliding order and sliding accuracy in sliding mode control. Int. J. Control, 58: 1247–1263.
22. G. Bartolini, A. Ferrara, E. Usai, and V. Utkin, 2000. On multi-input chattering-free second-order sliding mode control. IEEE Trans. Automat. Control, 45 (9): 1711–1717.
23. M. Ouali, M.B.A. Kamoun, 1997. Field-oriented control induction machine and control by sliding mode. Simulation Practice and Theory 5: 121-136.
24. A. Mansouri, M. Chenafa, A. Bouhenna, E. Etien, 2004. Powerful Nonlinear Observer Associated With Field-Oriented Control Of an Induction Motor. Int. J. Appl. Math. Comput. Sci, 14 (2): 209–220.
25. Franck Morand, 2005. Techniques d'observation sans capteur de vitesse en vue de la commande des machines asynchrones. Thesis, National School Of Applied Sciences Of Lyon. 2005.
26. Utkin, V.I., 1977. Variable structure systems with sliding mode, IEEE Trans., AC-22: 212-222.
27. T.-L. Chern and Y.-C. Wu, 1993. Design of brushless DC position servo systems using integral variable structure approach. IEE Proceedings-B, 140 (1): 27-34.
28. K. JAMOUSSI, M. OUALI, H. CHARRADI, 2006. Reconstructeur d'état d'un moteur asynchrone par DSP. Proceeding of the seventh International Conference on Science and Techniques of Automatic control STA, 2006 RME-028, pp: 1:11.