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Performance Analysis of Adaptive Fuzzy Sliding Mode for Nonlinear Control of the Doubly Fed Induction Motor

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Article Info	ABSTRACT
	In this article, we propose a contribution to the control of a doubly fed induction motor by sliding mode with adaptive fuzzy logic. The technique of vector-control by classical field oriented applied to the doubly fed induction motor (DFIM) with mechanical sensors made it possible to have performances comparable with that of the direct current motor. However, it very sensitive to the parametric variations of the machine. The regulation speed by a classical regulator (PI) presents disadvantages: Poor robustness against parametric uncertainties of modeling and no the considering of the disturbances and little degree of freedom for the regulation. Because this effect, several robust controls were proposed in the technical literature to ensure the decoupling of the currents of the DFIM in a reference (d, q) leading to calculate simplified correctors. Among them, the variable structure control by sliding mode. It uses algorithms of regulations which ensure the robustness of the behavior of the process compared to the parametric variations and disturbances. Also, the impact of regulators based on artificial intelligence techniques such as adaptive fuzzy sliding mode controller are studied. In terms of results obtained, good dynamic performance and robustness with respect to load disturbances and parametric variation has been observed.

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1.   INTRODUCTION

In the field of high-power drives (such as the rolling mill for example), there is a new and original solution, using an alternative machine operating in a mode. This is the doubly fed induction motor DFIM, [1], [2] where the stator is fed by a fixed network and the rotor by variable power supply which can be a voltage source or a power source, [3].

DFIM finds its application in high power drives, it is characterized by its robustness, its longevity and a wider range of speed variation (hypo-synchronous, synchronous and hyper-synchronous regime) However, the asynchronous machine has a major disadvantage: the structure dynamic is strongly non-linear and the existence of a strong coupling between the torque and the flux, which complicates its control [4], [5].

In order to obtain a dual power asynchronous machine whose performances are like a DC machine, it is necessary to ensure the decoupling between the flux and the electromagnetic torque. This is the idea of the appearance of the vector control technique, or the control by flux orientation, [6]. This technique was proposed in 1973 by Blaschke and Hasse. The purpose of this technique is to manage the asynchronous machine as an independent excitation DC machine where there is a natural decoupling between the magnitude controlling the flux (the excitation current) and that related to the torque (the armature current) [7], [8].

Conventional PI or PID control techniques cover a wide range in industrial applications. These are linear control techniques and have the advantage of simplicity of implementation and ease of synthesis. Over time, its applications will be inefficient, especially if the processes to be controlled have complex and non-linear structures. Finding the substitute for these classic techniques is the concern of many researchers. Because the latter involves a compromise between the robustness of one side and the cost of another side [9], [10].

Sliding mode control, which was developed in the Soviet Union there are more than 30 years old, is a nonlinear control technique. It uses setting algorithms which ensure the robustness of the behavior of the process in relation to the variations parametric and disturbances. It has several advantages such as: Robustness, speed of response without exceeding and good efficiency at parametric and modeling uncertainties. However residual vibrations at high frequencies inherent or chattering to order remain considerable inconvenience, [11]-[13].

In order to eliminate the chattering [14], as it can damage the actuators by oscillations too frequent and affect the operation and performance of the system, it is proposed the application of fuzzy logic in sliding mode, [15], [16] for adjusting the speed of the DFIM.

This paper will be organized as follows. Section II dynamic model of DFIM and stator flux orientation will be reported; principle of Sliding mode control will be given in Section III, as well as the proposed solution will be presented. In Section IV, results of simulation tests are reported. Concluding remarks will be given in section V.

2.   DOUBLY FED INDUCTION MACHINE MODEL

By referring to a rotating reference frame, denoted by the superscript (d, q), the dynamic model of a DFIM can be expressed by, [1]:

$$\begin{cases} \dfrac{di_{sd}}{dt} = -a_1 i_{sd} + (a\omega + \omega_s)i_{sq} + a_3 i_{rd} + a_5 \omega\, i_{rq} + b_1 v_{sd} - b_3 v_{rd} \\[2mm] \dfrac{di_{sq}}{dt} = -(a\omega + \omega_s)i_{sd} - a_1 i_{sq} - a_5 \omega\, i_{rd} + a_3 i_{rq} + b_1 v_{sq} - b_3 v_{rq} \\[2mm] \dfrac{di_{rd}}{dt} = a_4 i_{sd} - a_6 \omega\, i_{sq} - a_5 \omega\, i_{rd} + \left(\omega_s - \dfrac{\omega}{\sigma}\right)i_{rq} - b_3 v_{sd} + b_2 v_{rd} \\[2mm] \dfrac{di_{rq}}{dt} = -a_6 \omega\, i_{sd} - a_4 i_{sq} - \left(\omega_s - \dfrac{\omega}{\sigma}\right)i_{rd} - a_5 \omega\, i_{rq} - b_3 v_{sq} + b_2 v_{rq} \\[2mm] \dfrac{d\Omega}{dt} = \dfrac{1}{J} T_{em} - \dfrac{f}{J}\Omega - \dfrac{1}{J} T_r \end{cases} \tag{1}$$

where:

$$a = \frac{1-\sigma}{\sigma}, \; a_1 = \frac{R_s}{\sigma L_s}, \; a_2 = \frac{R_r}{\sigma L_r}, \; a_3 = \frac{R_r M}{\sigma L_r L_s}, \; a_4 = \frac{R_s M}{\sigma L_s L_r}, \; a_5 = \frac{M}{\sigma L_s}, \; a_6 = \frac{M}{\sigma L_r},$$

$$b_1 = \frac{1}{\sigma L_s}, \; b_2 = \frac{1}{\sigma L_r}, \; b_3 = \frac{M}{\sigma L_s L_r}, \; \sigma = 1 - \frac{M^2}{L_s L_r}$$

To simplify the control, it is necessary to make a judicious choice of reference. For this, one places oneself in a reference frame (d-q) linked to the rotating field with an orientation of the stator flux (the axis d aligned with the direction of the stator flux), as the following Figure 1 shows, [17].
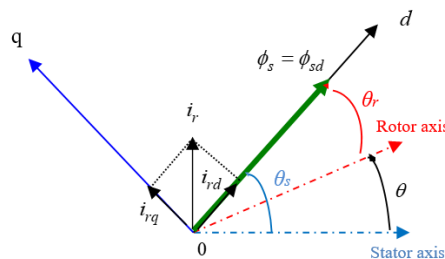

Figure 1. Stator field orientation on the d-axis

We obtain:

$$\phi_{sd} = \phi_s \text{ and } \phi_{sq} = 0 \tag{2}$$

The fluxes are given by:

$$\begin{cases} \phi_{sd} = L_s i_{sd} + M i_{rd} \\ \phi_{sq} = L_s i_{sq} + M i_{rq} \\ \phi_{rd} = L_r i_{rd} + M i_{sd} \\ \phi_{rq} = L_r i_{rq} + M i_{sq} \end{cases} \tag{3}$$

We obtain:

$$\begin{cases} i_{sd} = \dfrac{1}{L_s}(\phi_{sd} - M.i_{rd}) \\ i_{sq} = \dfrac{1}{L_s}(\phi_{sq} - M.i_{rq}) \end{cases} \tag{4}$$

The electromagnetic torque is expressed by:

$$T_{em} = \frac{pM}{L_s}(\phi_{sq} i_{rd} - \phi_{sd} i_{rq}) \tag{5}$$

Replacing (2) in (3) and (5) we find:

$$T_{em} = \frac{pM}{L_s}(-\phi_s i_{rq}) = -\frac{pM}{L_s}\phi_s i_{rq} \tag{6}$$

and

$$\begin{cases} \phi_{sq} = 0 \Rightarrow i_{sq} = -\dfrac{M}{L_s} i_{rq} \\ i_{sd} = 0 \\ i_{rd} = \dfrac{\phi_s^*}{M} \end{cases} \tag{7}$$

from (6)

$$i_{rq} = -\frac{L_s}{pM}\frac{T_{em}^*}{\phi_s^*} \tag{8}$$

$$\frac{d\theta_s}{dt} = \omega_s = \frac{\left(\dfrac{R_s.M}{L_s} i_{rq} + V_{sq}\right)}{\phi_s^*} \tag{9}$$

after arrangement we will have the following system:

$$\begin{aligned} \frac{di_{rd}}{dt} &= -\frac{1}{\sigma}\left(\frac{1}{T_r} + \frac{M^2}{L_s T_s L_r}\right) i_{rd} - \frac{M}{\sigma L_s L_r} V_{sd} + \frac{M}{\sigma L_s L_r T_s}\phi_{sd} + (\omega_s - \omega)i_{rq} + \frac{1}{\sigma L_r}v_{rd} \\ \frac{di_{rq}}{dt} &= -\frac{1}{\sigma}\left(\frac{1}{T_r} + \frac{M^2}{L_s T_s L_r}\right) i_{rq} - \frac{M}{\sigma L_s L_r} V_{sq} + \frac{M}{\sigma L_s L_r}\omega\,\phi_{sd} + (\omega_s - \omega)i_{rd} + \frac{1}{\sigma L_r}v_{rq} \\ \frac{d\phi_{sd}}{dt} &= V_{sd} + \frac{M}{T_s} i_{rd} - \frac{1}{T_s}\phi_{sd} \\ \frac{d\phi_{sq}}{dt} &= V_{sq} + \frac{M}{T_s} i_{rq} - \omega_s \phi_{sd} \\ \frac{d\Omega}{dt} &= -\frac{pM}{JL_s}(\phi_{sd} i_{rq}) - \frac{f}{J}\Omega - \frac{1}{J}T_r \end{aligned} \tag{10}$$

with: $T_r = \dfrac{L_r}{R_r}$; $T_s = \dfrac{L_s}{R_s}$

The flux estimator can be obtained by the following (11):

$$\phi_{sd} = L_s i_{sd} + M i_{rd}$$
$$\phi_{sq} = L_s i_{sq} + M i_{rq}$$

(11)

The position stator flux is calculated by the following equations: $\theta_r = \theta_s - \theta$

where $\theta_s = \int \omega_s dt$, $\theta = \int \omega dt$, $\omega = P\Omega$. θ_s is the electrical stator position, θ is the electrical rotor position.

3.   PRINCIPLE OF SLIDING MODE CONTROL

A Sliding Mode Controller (SMC) is a Variable Structure Controller (VSC). Basically, a VSC includes several different continuous functions that can map plant state to a control surface. As a result, the sliding mode control is based on three steps [18].

3.1. Choice of the switching surface

J. Slotine proposes a form of general equation to determine the sliding surface

$$S(x) = \left(\frac{d}{dt} + \lambda\right)^{n-1} e(x)$$

(12)

$$e(x) = x_{ref} - x$$

(13)

$e(x)$: denotes the error of the controlled greatness;

λ : Positive coefficient;

n : Relative degree;

x_{ref} : Reference greatness.

3.2. The condition of convergence

The condition of convergence is defined by the equation of Lyapunov.

$$S(x).\dot{S}(x) < 0$$

(14)

3.3. Control Calculation

The control algorithm includes two terms, the first for the exact linearization, the second discontinuous one for the system stability.

$$u = u^{eq} + u^{n}$$

(15)

u^{eq} : corresponds to the equivalent order suggested by Utkin (1992). It is calculated starting from the expression:

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

(16)

u^{n} is given to guarantee the attractivity of the variable to be controlled towards the commutation surface.

The simplest equation is the form of relay.

$$u^{n} = k \, sat \, S(x)$$

(17)

k is the controller gain.

3.4. Speed control

The speed error is defined by:

$$e = \Omega_{ref} - \Omega \tag{18}$$

for $n = 1$, the speed control manifold equation can be obtained by:

$$S(\Omega) = e = \Omega_{ref} - \Omega \tag{19}$$

$$\dot{S}(\Omega) = \dot{\Omega}_{ref} - \dot{\Omega} \tag{20}$$

substituting the expression of $\dot{\Omega}$ in (10) and in (20), we obtain:

$$\dot{S}(\Omega) = \dot{\Omega}_{ref} - \left(-\frac{pM}{JL_s}(\phi_{sd}i_{rq}) - \frac{f}{J}\Omega - \frac{1}{J}T_r \right) \tag{21}$$

we take:

$$i_{rq} = i_{rq}^{eq} + i_{rq}^{n} \tag{22}$$

during the sliding mode and in permanent regime, we have:

$$S(\Omega) = 0, \ \dot{S}(\Omega) = 0, \ i_{rq}^{n} = 0$$

the equivalent control is:

$$i_{rq}^{eq} = -\frac{JL_s}{PM\phi_{sd}}\left(\dot{\Omega}_{ref} + \frac{C_r}{J} + \frac{f}{J}\Omega \right) \tag{23}$$

where the correction factor is given by:

$$i_{rq}^{n} = K_{i_{rq}} sat(S(\Omega)) \tag{24}$$

In our proposal which allows the combination between the fuzzy logic and the control by sliding mode, we call the controller resulting from this combination: controller by fuzzy sliding mode (FSMC), it has the same control law as the SMC set to apart from the parameters k and ξ the component u^n that will be adapted by a system to a fuzzy inference.

The key idea of this combination is inspired by the fact that in the ideal case and when $S(x)$ is far from the sliding surface, parameters k and ξ must have a large enough value and when $S(x)$ is near of the sliding surface, adjusted parameters k and ξ must have a small value.

In this voice one can propose an adaptation of these parameters by means of a system with a fuzzy inference while keeping the same laws of control of the controller SMC.
$S(x)$ in our case represents the input space of the fuzzy system composed of two terms S and $\dot{S}$, the gain adjustment k and the parameter ξ are written in the form of the following fuzzy rule:

If S (t) is A$_j$ Then k and ξ are B$_j$ (25)

where j = 1, ..., m and m represent the total number of rules.

The terms k and ξ are therefore adapted by a fuzzy adapter having two inputs $"\varepsilon"$ and $"\Delta\varepsilon"$ of three membership functions and an $'\alpha'$ output of nine membership functions which are represented in Figure 2 and Figure 3 respectively. A typical rule reads as follows, as shown in Table 1.
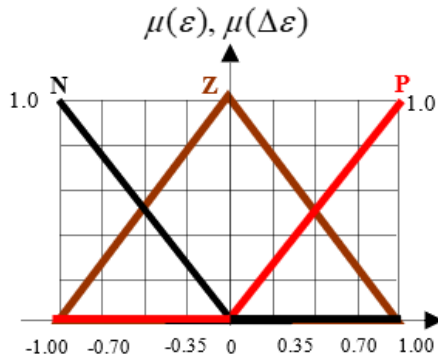
Figure 2. The inputs membership
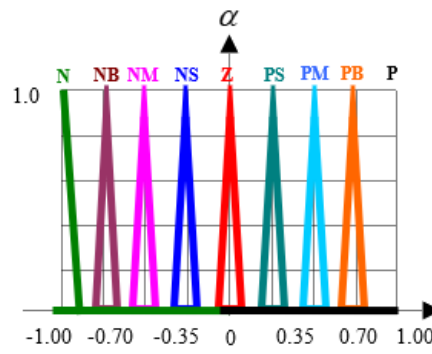functions of the AFSMC

Figure 3. The output membership
functions of the AFSMC

Table 1. A typical rule reads as follows:

Symbol	Mean	Symbol	Mean
N	Negative	P	Positive
Z	Zero	NB	Negative Big
NS	Negative Small	NM	Negative Medium
PS	Positive Small	PM	Positive Medium
PB Positive Big			

therefore, the discontinuous control law of (24) becomes:

$$i_{rqn}^{fl} = k_{\Omega}^{fl} \, sat\left(s(\Omega) \Big/ \xi_{\Omega}^{fl} \right) \tag{26}$$

with

$$\begin{cases} k_{\Omega}^{fl} = \alpha \, k_{\Omega} \\ \xi_{\Omega}^{fl} = \alpha \, \xi_{\Omega} \end{cases}$$

For the control law of (26), we will see in a transient state a series of values of k_{Ω}^{fl} and ξ_{Ω}^{fl}, which gives an infinite number of $sat\left(s(\Omega) \Big/ \xi_{\Omega}^{fl} \right)$ functions. In steady state, the magnitude α takes the value '1'. The result of defuzzifying the command for an entry $S(x)$ is presented as shown in Figure 4.
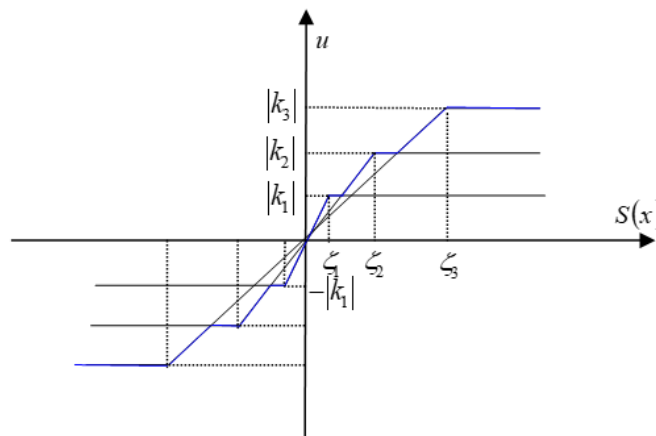

Figure 4. Control signal of the fuzzy-sliding controller

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The rules governing the gain adaptation law and the parameter are presented as follows:

R_1 : if	S	is	P and $\dot{S}$ is	P then	α	is	N ;
R_2 : if	S	is	P and $\dot{S}$ is	Z then	α	is	NB ;
R_3 : if	S	is	P and $\dot{S}$ is	N then	α	is	NM ;
R_4 : if	S	is	Z and $\dot{S}$ is	P then	α	is	NS ;
R_5 : if	S	is	Z and $\dot{S}$ is	Z then	α	is	ZE ;
R_6 : if	S	is	Z and $\dot{S}$ is	N then	α	is	PS ;
R_7 : if	S	is	N and $\dot{S}$ is	P then	α	is	PM ;
R_8 : if	S	is	N and $\dot{S}$ is	Z then	α	is	PB ;
R_9 : if	S	is	N and $\dot{S}$ is	N then	α	is	P

3.5. Rotor direct current control and limitation

In order to limit all possible overshoot of the current i_{rd}, we add a limiter of current defined by:

$$i_{rd}^{\cdot \lim} = i_{rd}^{\cdot \max} sat(i_{rd}) \tag{27}$$

The direct current error is defined by:

$$e = i_{rd}^{\cdot \lim} - i_{rd} \tag{28}$$

for $n = 1$, the direct current control manifold equation can be obtained by:

$$S(i_{rd}) = i_{rd}^{\cdot \lim} - i_{rd} \tag{29}$$

$$\dot{S}(i_{rd}) = i_{rd}^{\cdot \lim} - i_{rd}^{\cdot} \tag{30}$$

Substituting the expression of $i_{rd}^{\cdot}$ in (10) and in (30), we obtain:

$$\dot{S}(i_{rd}) = i_{rd}^{\cdot \lim} - \left(-\frac{1}{\sigma}\left(\frac{1}{T_r} + \frac{M^2}{L_s.T_s.L_r}\right)i_{rd} - \frac{M}{\sigma L_r L_s}V_{sd} + \frac{M}{\sigma L_r L_s T_s}\phi_{sd} + (\omega_s - \omega)i_{rq} + \frac{1}{\sigma L_r}V_{rd} \right) \tag{31}$$

The control voltage V_{rd}^{ref} is defined by

$$V_{rd}^{ref} = V_{rd}^{eq} + V_{rd}^{n} \tag{32}$$

during the sliding mode and in permanent regime, we have:

$$S(i_{rd}) = 0 \ , \ \dot{S}(i_{rd}) = 0 \ , \ V_{rd}^{n} = 0$$

where the equivalent control is:

$$V_{rd}^{eq} = \left(i_{rd}^{\cdot \lim} + \frac{1}{\sigma}\left(\frac{1}{T_r} + \frac{M^2}{L_s.T_s.L_r}\right)i_{rd} + \frac{M}{\sigma L_r L_s}V_{sd} - \frac{M}{\sigma L_r L_s T_s}\phi_{sd} - (\omega_s - \omega)i_{rq} \right)\sigma L_r \tag{33}$$

substituting $i_{rd} = \dfrac{\phi_s^*}{M}$ in (23)

$$V_{rd}^{eq} = \left(\dot{i}_{rd}^{\lim} + \frac{1}{\sigma T_r} i_{rd} - \frac{M}{\sigma L_r L_s} V_{sd} - (\omega_s - \omega) i_{rq} \right) \sigma L_r \tag{34}$$

therefore, the correction factor is given by:

$$V_{rd}^n = K_{V_{rd}} sat(S(i_{rd})) \tag{35}$$

where $K_{V_{rd}}$:positive constant

3.6. Rotor quadrature current control and limitation
In order to limit all possible overshoot of the current i_{rd}, we add a limiter of current defined by:

$$i_{rq}^{\lim} = i_{rq}^{\max} sat(i_{rq}) \tag{36}$$

the direct current error is defined by:

$$e = i_{rq}^{\lim} - i_{rq} \tag{37}$$

for $n = 1$, the direct current control manifold equation can be obtained by:

$$S(i_{rq}) = i_{rq}^{\lim} - i_{rq} \tag{38}$$

$$\dot{S}(i_{rq}) = \dot{i}_{rq}^{\lim} - \dot{i}_{rq} \tag{39}$$

Substituting the expression of $\dot{i}_{rq}$ equation (10) in equation (39), we obtain:

$$\dot{S}(i_{rq}) = \dot{i}_{rq}^{\lim} - \left(-\frac{1}{\sigma} \left(\frac{1}{T_r} + \frac{M^2}{L_s.T_s.L_r} \right) i_{rq} - \frac{M}{\sigma L_r L_s} V_{sq} + \frac{M}{\sigma L_r L_s} \omega_s \phi_{sd} - (\omega_s - \omega) i_{rd} + \frac{1}{\sigma L_r} V_{rq} \right) \tag{40}$$

the control voltage V_{rq}^{ref} is defined by

$$V_{rq}^{ref} = V_{rq}^{eq} + V_{rq}^n \tag{41}$$

during the sliding mode and in permanent regime, we have:

$$S(i_{rq}) = 0, \ \dot{S}(i_{rq}) = 0, \ V_{rq}^n = 0$$

where the equivalent control is:

$$V_{rq}^{eq} = \left(\dot{i}_{rq}^{\lim} + \frac{1}{\sigma} \left(\frac{1}{T_r} + \frac{M^2}{L_s.T_s.L_r} \right) i_{rq} + \frac{M}{\sigma L_r L_s} V_{sq} - \frac{M}{\sigma L_r L_s} \omega_s \phi_{sd} + (\omega_s - \omega) i_{rq} \right) \sigma L_r \tag{42}$$

therefore, the correction factor is given by:

$$V_{rq}^n = K_{V_{rq}} sat(S(i_{rq})) \tag{43}$$

where $K_{V_{rq}}$:positive constant

3.7. Stator flux control

In the proposed control, the manifold equation can be obtained by:

$$S(\phi_{sd}) = \phi_{sd}^{ref} - \phi_{sd} \tag{44}$$

$$\dot{S}(\phi_{sd}) = \dot{\phi}_{sd}^{ref} - \dot{\phi}_{sd} \tag{45}$$

Substituting the expression of ϕ_{sd} equation (10) in equation (45), we obtain:

$$\dot{S}(\phi_{sd}) = \dot{\phi}_{sd}^{ref} - \left(V_{sd} + \frac{M}{T_s} i_{rd} - \frac{1}{T_s} \phi_{sd} \right) \tag{46}$$

the control voltage i_{rd} is defined by

$$i_{rd} = i_{rd}^{eq} + i_{rd}^{n} \tag{47}$$

during the sliding mode and in permanent regime, we have:

$$S(\phi_{sd}) = 0, \ \dot{S}(\phi_{sd}) = 0, \ i_{rd}^{n} = 0$$

The equivalent control is:

$$i_{rd}^{eq} = \left(\dot{\phi}_{sd}^{ref} - V_{sd} + \frac{1}{T_s} \phi_{sd} \right) \frac{M}{T_s} \tag{48}$$

where the correction factor is given by:

$$i_{rd}^{n} = K_{i_{rd}} sat(S(i_{rd})) \tag{49}$$

where $K_{i_{rd}}$:positive constant

4. RESULTS AND ANALYSIS

To evaluate the performance of the proposed method, the model for the control law applied to the DFIM Figure 5 has been implemented in the Matlab / Simulink software.

The simulation results illustrate the dynamic responses of the rotor speed, the electromagnetic torque, the components of the stator fluxes and the rotor current.

The motor is operated at Ω_{ref}=150 rad/s under no load and a load disturbance torque (10 N.m) is suddenly applied at t=1s and eliminated at t=2s (-10 N.m), To show the effect of the parameters uncertainties, we have simulated the system with +50% of value of the rotor resistance and compared to nominal value.

This Figure 6 shows the dynamic response of the speed. The latter reaches its reference in a response time of 0.13 s. The introduction of charge has no influence on the evolution (stability) of the speed, which shows the robustness of the proposed control against these disturbances. It is also noted that the speed follows its setpoint rapidly during the reverse direction of rotation (identical to the response time at startup).

It can also be seen that increasing the rotor resistance R_r by + 50% of its nominal value does not cause any undesirable effects.
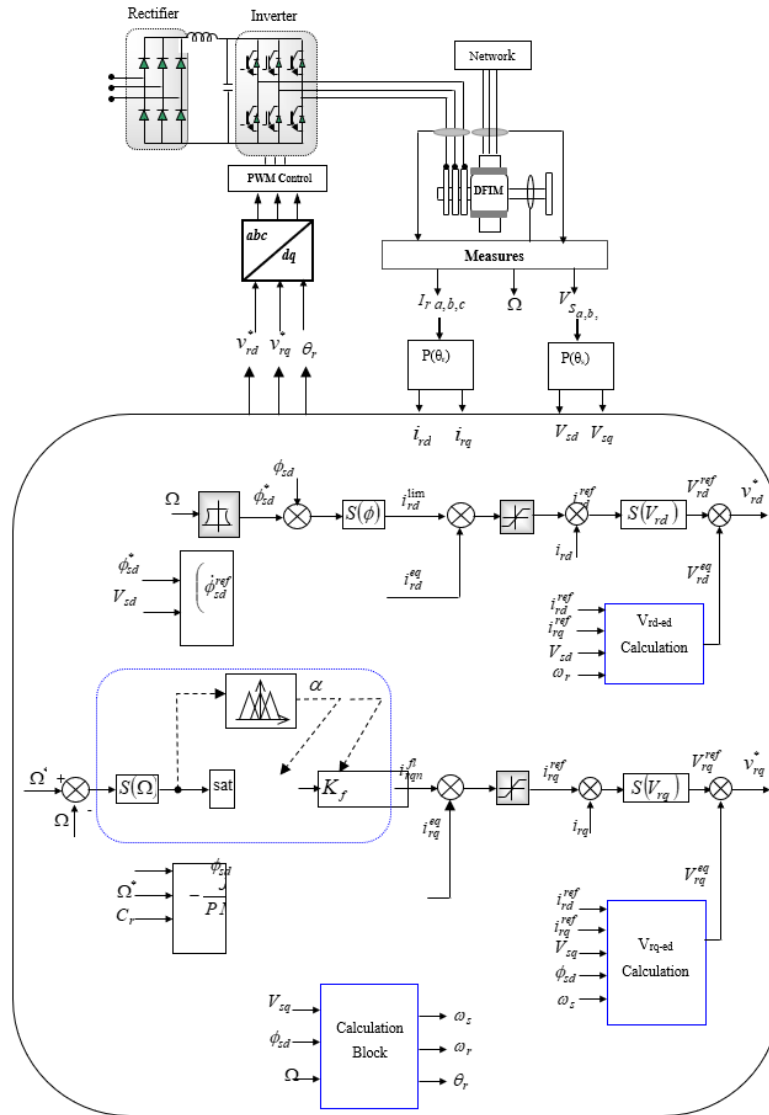
Figure 5. Non-Linear Control scheme for DFIM using the Adaptive Fuzzy Sliding Mode controller
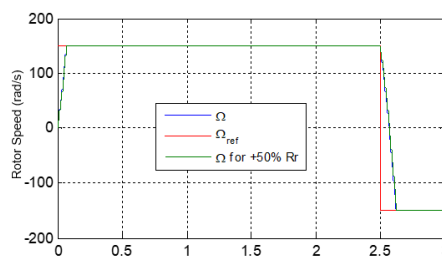

Figure 6. The performance of the rotor speed for DFIM using the Adaptive Fuzzy Sliding Mode controller

The simulation reported in Figure 7 shows that the direct component of the stator flux stabilizes at the value of the steady state reference stator flux while its quadrature component stabilizes at zero ($\phi_{sd} = \phi_{ref}$, $\phi_{sq} = 0$).

During the rotation inversion the change of the direction of the torque does not degrade the orientation of the flux so the introduction of charge has no influence, and it is clearly seen that the decoupling is not influenced by the variation of R_r.

Figure 8 illustrates the dynamic response of the electromagnetic torque for two rotor resistance values (R_{rn} and + 50% of R_r). Given this result, we note that this command gives good performance. It is noted that the regulation of the two rotor currents by sliding mode Figure 9 gives faster responses during a vacuum start and more robust during the variation of the load and the rotor resistance.
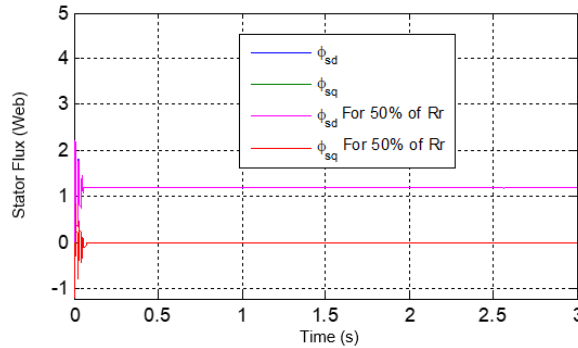
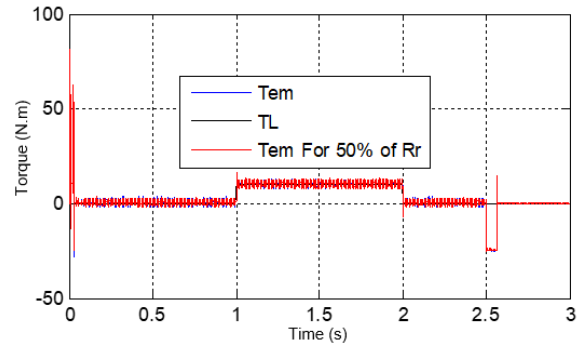Figure 7. The performance of the stator flux for DFIM Sliding Mode controller

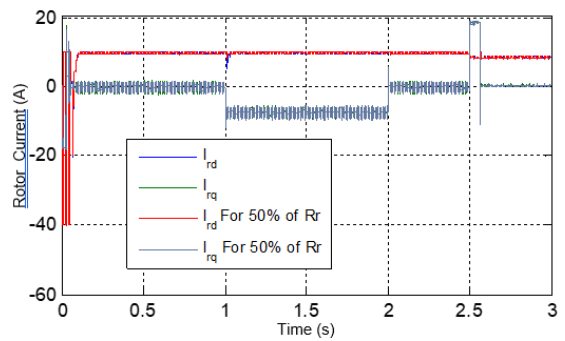Figure 8. The performance of the torque for DFIM

Figure 9. The performance of the rotor current for DFIM using Sliding Mode controller

From the results obtained, all the tests applied to this control plus the variation of the rotor resistance does not cause any undesirable effect on all the dynamic responses, and this shows the robustness of the nonlinear control applied to the DFIM.

Finally, and after this detailed study, we note a significant improvement in the static and dynamic performance of the DFIM, which is proving the efficiency and robustness of the developed Adaptive Fuzzy Sliding Mode controller and the nonlinear control chosen compared to the classic PI controller developed in [19], [20]

5.   CONCLUSION

This paper was the subject of the application of the sliding mode controller in the following control loops: stator flux and rotor currents on the Doubly Fed Induction Motor (DFIM) and it is applied Adaptive Fuzzy Sliding Mode controller only in the speed loop.

in this context, we first presented a theoretical reminder on the sliding mode control of variable structure systems, then we approached the adaptive fuzzy control algorithm with its different steps, then applied it on the regulation of DFIM, and finally simulation results are presented and commented.

It can be concluded from these results that, the field-oriented control using a non-linear Fuzzy-sliding mode is more robust than that of conventional regulator-controlled vector control (IP or PI). The comparative study in the test results by speed reference variation and rotor resistance variation shows that the robustness quality of the proposed controller.

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