Contribution to Study Performance of the Induction Motor by Sliding Mode Control and Field Oriented Control

Oukaci Assia, Toufouti Riad and Dib Djalel

Abstract The induction motor squirrel cage that is deemed by its strength, high torque mass, robustness, and its relatively low cost ... etc., meanwhile, it benefited from the support of industry since its invention (invention by Tesla the late nineteenth century). Unfortunately, these advantages are accompanied by a high complexity of the physical interactions between the stator and the rotor. Therefore, dynamic control requires complex control algorithms in contrast to its structural simplicity. In recent decades, many techniques of control of the induction machine, such as technical oriented control or Field Oriented control, have emerged and are currently used to enjoy the benefits of the asynchronous machine for applications where variable speed is essential. The high operating control of the induction machine began with the invention of the oriented vector control in the late 60s flux. Before that time control of the induction machine was limited to scalar commands. This operating control does not provide a decoupling between the flux and torque. To illustrate this, the torque of a cage induction motor has to be increased by increasing the slip, the flux is affected by a decrease; therefore the torque control is dependent of the stream, for this the inherent coupling between these two variables makes conventional techniques less efficient. To solve these problems this paper seeks to analyze dynamical performances and sensitivity to induction motor parameter changes, two techniques are applied Sliding Mode Control and Field Oriented Control. For this, this design on the basis of some simulations results is illustrated with different functions in order to illustrate its efficiency and make comparison between the two techniques; Numerical simulations are presented to validate the proposed methods. The objective of this paper is to

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guarantee the desired performance of the induction motor, robust to the parameters variations, disturbances, and reach the speed of rotation at the speed desired in a minimum response time.

Nomenculture

r, s : Subscripts stand for rotor and stator;
R_r, R_s : Rotor and stator resistances;
L_r, L_s, L_m : Rotor, stator and mutual inductances;
C_{em} : Electromagnetic torque;
C_r : Load torque;
σ : Total leakage coefficient;
J : Moment of inertia;
v, i : Voltage and current;
ϕ : Flux linkage;
ω_r : Electrical angular rotor speed;
ω_s : Synchronously rotating angular speed;
p : Number of poles pair.

1 Introduction

Currently, induction motors (IM) are widely used in many industrial applications, including transportation, conveyor systems, actuators, material han-dling, pumping of liquid metal, and others, with satisfactory performance (Boucheta et al. 2012). The electromechanical systems of this motor are suitable for a large spectrum of industrial applications (Isidori 1995). However, induction motors are multivariable nonlinear and strongly coupled time-varying systems, mainly, in variable speed applications. However, its dynamic control requires complex control algorithms, facing its structural simplicity, since there is a complex coupling between the input variables, output variables and the internal variables of the machine (Leonhard 1994; Meziane et al. 2008; Maher 2012).

Induction motors are suitable electromechanical systems for a large spectrum of industrial applications. However, induction motors are multivariable nonlinear and strongly coupled time-varying systems, mainly, in variable speed applications (Leonhard 1994; Hautier and Caron 1995). However, its dynamic control requires complex control algorithms, facing its structural simplicity, since there is a complex coupling between the input variables, output variables and the internal variables of the machine (Isidori 1995; Meziane et al. 2008; Maher 2012).

So with the invention of power electronics, and advances in computing, has made a radical revolution. Its goal is to develop control strategies for induction motors. The design of suitable control algorithms for those motors (IM) has been widely investigated for more than two decades. Since the beginning of field oriented control (FOC) of AC drives works like a separately excited DC motor and it was proposed by Blaschke (Direct FOC) and Hasse (Indirect FOC) in early 1970s, (Ramesh et al. 2013), seen as a viable replacement of the traditional DC drives, several techniques from linear control theory have been used in the different control loops of the FOC scheme, such as Proportional Integral (PI) regulators, and exact feedback linearization (Ouhrouche and Volat 2000; Duarte-Mermoud and Travieso-Torres 2012). Due to their linear characteristics, these techniques do not guarantee suitable machine operation for the whole operation range, and do not consider the parameter variations of the motor-load set. The field oriented control technique has a major disadvantage, such as; requirement of co-ordinate transformations, current controllers, sensitive to parameter variations. This drawbacks of FOC schemes are minimized with the new control strategy i.e., DTFC scheme, which is introduced by Isao Takahashi and Toshihiko Noguchi, in the mid 1980s, (Ramesh et al. 2013).

Direct torque and flux control of an IM is requires the rotor shaft angular position information. The rotor shaft position can be measured through either speed sensors (i.e., speed encoder) or from an estimator/observer using current and voltage signals and information of the IM parameters. The use of speed encoder is associated with some drawbacks, such as, requirement of shaft extension, reduction of mechanical robustness of the motor drive, reduces the drive reliability and not suitable for hostile environments, and also costlier. These drawbacks have made speed sensorless direct torque and flux controlled IM very attractive over the conventional direct torque and flux control (DTFC) drive. There are some applications for sensorless drive, where there is no sufficient space to put the speed sensor or the nature of the environment does not allow the use of any additional rotor speed sensors, (Ramesh et al. 2013).

Over the past years, several schemes have proposed for rotor speed estimation in the sensorless vector controlled IMs, (Caruana et al. 2006; Kojabadi 2005). They are: (i) signal injection based method (Caruana et al. 2006), (ii) state observer based method (Rojas et al. 2004), and (iii) model based method (Kojabadi 2005). The signal injection method is suffers from computational complexity and requirement of external hardware for signal injection, (Ramesh et al. 2013). Among these methods, the sliding mode technique is one of the nonlinear control techniques has also been proposed to solve the problems mentioned above (Rao et al. 2009; Saiad 2012).

Much research has been done in recent years to apply various approaches to attenuate the effect of uncertainties. On the basic aspect, the conventional proportional-integral-derivative (PID) controllers are widely used in industry due to their simple control structure, ease of design and low cost. (El-Sousy 2013; Holmes et al. 2012). However, the PID controller cannot provide perfect control performance if the controlled system is highly nonlinear and uncertain as in the case of IM. In addition, an objection to the real-time use of such control scheme is the lack of knowledge of uncertainties (El-Sousy 2013).

Due to the existence of nonlinearities, uncertainties, and disturbances, conventional linear control methods, including the PID control, cannot guarantee a sufficiently high performance for the IM servo drive system. To deal with these uncertainties, and to enhance the control performance, in recent years, many nonlinear control methods have been developed for the IM drive system, such as variable structure control (Liaw et al. 2001), adaptive and robust control (Xia et al. 2000; Ravi Teja and Chakraborty 2012), sliding mode control (SMC) (Li et al. 2005; Comanescu et al. 2008), higher-order sliding-mode control (Rashed et al. 2005; Traoré et al. 2008), fuzzy control, neural network control, wavelet neural net work control (Lin and Hsu 2002; Castillo-Toledo et al. 2008), hybrid control (Wai 2001; Wai et al. 2002), optimal control (Bottura et al. 2000; Attaianese and Timasso 2001), intelligent SMC (Wai et al. 2002; Zhu et al. 2011), supervisory control using genetic algorithm (Wai 2003; Su and Kung 2005) and so on. These approaches improve the control performance of the IM drive from different aspects. The sliding-mode control (SMC) is one of the effective nonlinear robust control approaches since it provides system dynamics with an invariance property to uncertainties once the system dynamics are controlled in the sliding mode (El-Sousy 2013).

Sliding-mode control has received much attention in the control of IM drives. It is well known that the major advantage of sliding mode control (SMC) systems is its insensitivity to parameter variations and external disturbance once the system trajectory reaches and stays on the sliding surface (Slotine and Li 1991; Veselic et al. 2010).

The robustness of the SMC is guaranteed usually by using a large switching control gain. This switching strategy often in the hitting control law (Slotine and Li 1991; Rao et al. 2009; Astrom and Wittenmark 1995; Corradini et al. 2012; El-Sousy 2013).

The sliding mode control (SMC) is a nonlinear control and based on the switching functions of state variables, used to create a variety or hyper sliding surface, whose purpose is to force the system dynamism to correspond with the defined by the equation of the hyper-surface. When the state is maintained on the hyper surface, the system is in sliding regime. Its dynamic is so insensitive to external disturbances and parametric conditions as sliding regime are carried out. In the synthesis of the control law by way of sliding, the sliding surface is defined as an independent and stable linear system. However, the dynamics imposed by such a system is slower than that imposed by a nonlinear system, hence the importance of using the latter type of systems to synthesize the sliding surface in some applications (Hautier and Caron 1995; Araujo and Freitas 2000; Aurora and Ferrara 2007; Bounar et al. 2012; Saiad 2012; Talhaoui et al. 2013).

1.1 Chapter Objectives

The main objective of this work is to improve the performance of the electric machine converter association. Indeed, a new technique for robust control by sliding mode is presented. The application example is given in this work is that the induction motor.

The bulk of this work thereafter is to arrive at clear a comparative study between conventional vector control orientation of the rotor flux and the new technique of sliding mode control. This leads to show fields and limits of use of each control, while by highlighting all the features that differentiate them both. For this, the proposed controls are applied to achieve a speed- and flux-tracking in a minimum response time objective under parameter uncertainties and disturbance of load thrust force

1.2 Chapter's Structure

The reminder of the current chapter is organized as follows: The first part focuses on the development of a mathematical model for an induction motor, the second part focuses on the development of the mathematical model for the field oriented control technique to induction motor, the third part focuses on the development of the sliding mode control to the induction motor, the last part of this chapter focuses on the performance analysis of theoretical results for both techniques to make a comparison between them. They have been validated by numerical simulations in Matlab/Simulink environment.

2 Nonlinear Induction Motor Model

Induction motor as various electric machines constitutes a theoretically interesting and practically important class of nonlinear dynamic systems. Induction motor is known as a complex nonlinear system in which time-varying parameters entail additional difficulty for induction motor system control, conditions monitoring and faults diagnosis (Leonhard 1994; Isidori 1995). Based on the fact that the nonlinear model of the induction motor system can be significantly simplified, if only one applies the d-q Park transformation (Appendix 1) (Jimoh et al. 2012), different structures of the nonlinear model are investigated. The choice of a model depends on measurement possibilities, selected state variables of the machine and the problem at hand. In this paper, the considered induction motor model has stator current, rotor flux and rotor angular velocity as selected state variables. The control inputs are the stator voltage and load torque. The available stator current measurements are the induction motor system outputs. The nonlinear state space model of the induction motor is expressed as the following (Hautier and Caron 1995; Meziane et al. 2008; Mira and Duarte-Mermoud 2009):

$$\begin{cases} \dot{x} = f(x) + g(x).v \\ y = h(x) \end{cases} \tag{1}$$

With:

$$v = \begin{bmatrix} V_{s\alpha} & V_{s\beta} \end{bmatrix}^T = \begin{bmatrix} U_1 & U_2 \end{bmatrix}^T$$
$$x = \begin{bmatrix} i_{s\alpha} & i_{s\beta} & \phi_{r\alpha} & \phi_{r\beta} \Omega_r \end{bmatrix}^T = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 x_5 \end{bmatrix}^T$$

Such as :
x : State vector.
v : Vector control.
y : Output selected.
$h(x)$: An analytic function.

$$f(x) = \begin{bmatrix} a_{11} \cdot x_1 + a_{13} \cdot x_3 + a_{14} \cdot x_4.x_5 \\ a_{11} \cdot x_2 - a_{14} \cdot x_3.x_5 + a_{13} \cdot x_4 \\ a_{31} \cdot x_1 + a_{33} \cdot x_3 + a_{34} \cdot x_4.x_5 \\ a_{31} \cdot x_2 - a_{34} \cdot x_3.x_5 \quad + a_{33} \cdot x_4 \\ \mu \cdot (x_2 \cdot x_3 - x_1 \cdot x_4) - \frac{c_r}{J} \end{bmatrix}; g(x) = \begin{bmatrix} b_{11} & 0 \\ 0 & b_{11} \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$

Such as:

$$a_{11} = -\left[\frac{1}{\sigma \cdot T_s} + \frac{1}{T_r} \cdot \left(\frac{1-\sigma}{\sigma}\right)\right], \quad a_{13} = \frac{1-\sigma}{\sigma} \cdot \frac{1}{M \cdot T_r}, \quad a_{14} = \frac{1}{M} \cdot \frac{1-\sigma}{\sigma} \cdot p,$$
$$a_{31} = \frac{M}{T_r}, \quad a_{33} = -\frac{1}{T_r}, \quad a_{34} = -p, \quad b_{11} = \frac{1}{L_s.\sigma}, \quad \sigma = 1 - \frac{M^2}{L_s.L_r}, \quad T_s = \frac{L_s}{R_s},$$
$$T_r = \frac{L_r}{R_r}, \quad \mu = \frac{p \cdot M}{J \cdot L_r}.$$

3 Principle of Field Oriented Control FOC

The basic idea of this method of control is to bring the behavior of the asynchronous machine similar to that of a DC machine with separate excitation or decoupling is natural. This method is based on transforming the electric machine variables to a repository that rotates with the rotor flux vector oriented (Blaschke 1977; Hautier and Caron 1995; Rong-Jong et al. 2005; Meziane et al. 2008; Bouchhida et al. 2012). Therefore, it can control the flux of the machine with the component I_{sd} of the stator current which is the equivalent of the inductor DC current machine. While, the component I_{sq} to control the armature current corresponding to the DC machine, electromagnetic torque (Mira and Duarte-Mermoud 2009). This is shown in Fig. 1

The relationship of the electromagnetic torque of the DC machine is given by, (Hautier and Caron 1995):

$$C_e = \kappa \cdot \phi \cdot i_a = K \cdot i_f \cdot i_a \tag{2}$$

With:
ϕ : Flux imposed by the excitation current i_f and i_a : Inductive current.

The inductive current i_a is the magnitude of the torque generator and the excitation current i_f is the magnitude of the flux generator. Thus, in a DC machine everything happens as if the control variables i_f and i_a are orthogonal. This means that the current

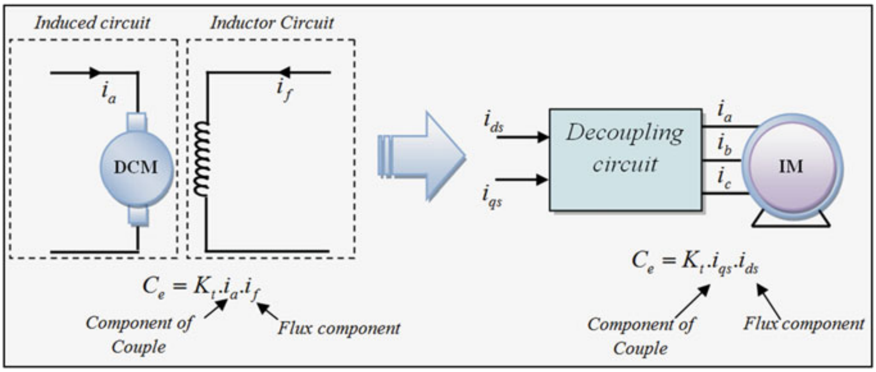

Fig. 1 Analogy induction machines with DC motor, (Hautier and Caron 1995; Meziane et al. 2008)

flux controlled by the i_f , and the current torque by the i_a . It is said that the armature and inductor are naturally decoupled.

Thus dissociates the stator current of the asynchronous machine in two components I_{sd} and I_{sq} in quadrature such that the current I_{sd} is oriented along the axis of the flux guide. A constant rotor flux, the couple then depend only aware I_{sq} (Hautier and Caron 1995; Chaigne and Etien 2005; Meziane et al. 2008).

3.1 Field Oriented Control Structure

The choice of the reference model of the induction machine from the template in the Park transformation is such that the axis (*d*) coincides with the desired direction of the flux (the rotor flux, the stator flux or air gap flux) as shown in the following Fig. 2, (Araujo and Freitas 2000; Mira and Duarte-Mermoud 2009; El-Sousy and Salem 2004):

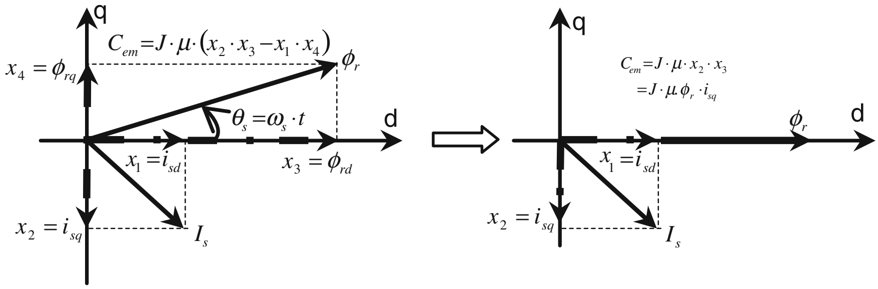

Fig. 2 Angular relations of current vectors, (Araujo and Freitas 2000)

There are three types of flux directing namely:

$$- \text{Rotor flux orientation} \quad : \phi_{\text{rd}} = \phi_r \quad \text{and} \quad \phi_{\text{rq}} = 0 \tag{3}$$

$$- \text{Stator flux orientation} \quad : \phi_{\text{sd}} = \phi_s \quad \text{and} \quad \phi_{\text{sq}} = 0 \tag{4}$$

$$- \text{Gap flux orientation} \quad : \phi_{\text{gd}} = \phi_g \quad \text{and} \quad \phi_{\text{gq}} = 0 \tag{5}$$

Control of the stator flux or flux gap is more complicated and does not provide full decoupling between torque and flux, for this vector control rotor flux orientation is the most used (Chaigne and Etien 2005; Saiad 2012).

The strategy of the vector control is to independently control the flux term and the current term to impose a couple. Keeping the control variables as $\left(V_{sd}, V_{sq}\right)$ and state variables such as stator currents $\left(i_{sd}, i_{sq}\right)$, the flux ϕ_r and the mechanical speed. When an electric motor drives a mechanical load it is essential to properly control the dynamics of it, to master the instantaneous torque of it. The thrust of the vector control is to have the asynchronous machine for a couple proportional to flux engine and a current like the DC machine. So, let's take the expression of the electromagnetic torque of the induction machine (Rao et al. 2009):

$$C_{em} = J \cdot \mu \cdot (x_2 \cdot x_3 - x_1 \cdot x_4) \tag{6}$$

In the reference dq which are projected the rotor flux and the stator current running at the speed of the rotating field, either in: $\theta_s = \omega_s \cdot t$

In order to have expression analogous to that electromagnetic torque of a DC motor the axis will be directed of the rotor flux, air gap torque of the expression becomes:

$$C_{em} = J \cdot \mu \cdot x_2 \cdot x_3 \tag{7}$$

By imposing the condition (3) to state the model of IM (1) supplied with voltage equations at the following reduced system is realized, (Boukettaya et al. 2008; Meziane et al. 2008; Bouchhida et al. 2012):

$$\begin{cases} i_{sq} = \frac{L_r}{P \cdot M} \cdot \frac{C_{em}^*}{\phi_r^*} \\ i_{sd} = \frac{1}{M} \cdot \left[T_r \cdot \frac{d\phi_r^*}{dt} + \phi_r^* \right] \\ \omega_r = \frac{M}{T_r} \cdot \frac{i_{sq}}{\phi_r^*} \\ \omega_s = \omega_m + \omega_r \\ v_{sd} = R_s \cdot \left[\sigma \cdot L_s \frac{di_{sd}}{dt} + i_{sd} + T_s \frac{(1-\sigma) \cdot \phi_r^*}{M} \text{-} \sigma \cdot T_s \cdot \omega_s \cdot i_{sq} \right] \\ v_{sq} = R_s \cdot \left[\sigma \cdot T_s \frac{di_{sq}}{dt} + i_{sq} + T_s \frac{(1-\sigma) \cdot \phi_r^*}{M} \text{-} \sigma \cdot T_s \cdot \omega_s \cdot i_{sd} \right] \end{cases} \tag{8}$$

The diagram of vector control with a flux model is given in Fig. 3:

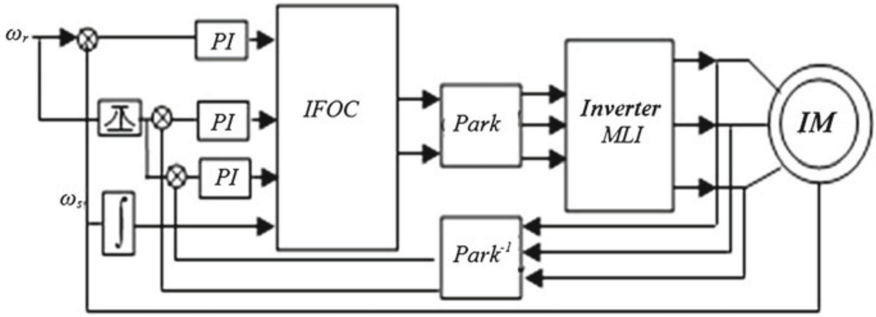

Fig. 3 Block diagram of indirect field oriented control structure (Meziane et al. 2008; Bouchhida et al. 2012)

4 General Concept of Sliding Mode Control

4.1 Condition Existence of Sliding Mode Control

The sliding mode exists when the switching takes place continuously between U_{max} and U_{min} . This is illustrated in Fig. 4 for the case of a control system of the second order with two state variables x_1 and x_2, (Wai 2003)

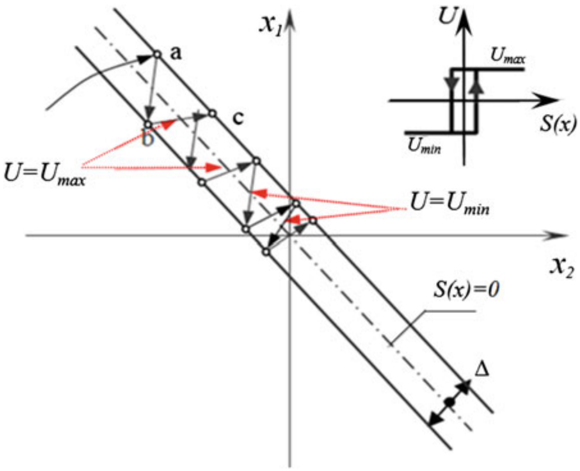

Fig. 4 Path of steady state sliding mode

4.2 Sliding Mode Control Design

Variable structure control (VSC) with sliding mode (SMC) is one of the effective nonlinear robust control approaches because it provides system dynamics with an invariance property to uncertainties once the system dynamics are controlled in the sliding mode. The first step of SMC design is to select a sliding surface that models the desired closed-loop performance in state variable space (Fig. 4). The control is then designed such that the system state trajectories are forced to the sliding surface and to stay on it. The system state trajectory in the period of time before reaching the sliding surface is called the reaching phase. Once the system trajectory reaches the sliding surface, it stays on it and slides along it toward the origin. The system trajectory sliding along the sliding surface toward the origin is the sliding mode. The insensitivity of the controlled system to uncertainties exists in the sliding mode, but not during the reaching phase. Thus, the system dynamic in the reaching phase continues to be influenced by uncertainties (Utkin 1993; Wai 2000; Ghanes and Zheng 2009; Boucheta et al. 2012).

4.2.1 The Choice of the Surface

The choice of the sliding surface for the necessary number and shape, depending on the application and purpose. In general, for a system defined by the state Eq. (1), choose “m” sliding surfaces for a vector of dimension “m”, (Dwards and Spurgeon 1998; Boucheta et al. 2012), with:

$$y(x) = \begin{bmatrix} y_1(x) \\ y_2(x) \end{bmatrix} = \begin{bmatrix} \Phi_r \\ \Omega_r \end{bmatrix} = \begin{bmatrix} \frac{1}{2}\left(x_3^2 + x_4^2\right) \\ x_5 \end{bmatrix} \tag{9}$$

The surface $S(x)$ represents the desired dynamic behavior of the system. J. J Slotine (Utkin 1993; Dwards and Spurgeon 1998; Wai 2000), proposes a form of general equation to determine the sliding surface which ensures the convergence of a variable towards its desired value x_{iref} .

If x_i a variable to controlled, associated with the following surface:

$$S_i(x_i) = \left(\frac{d}{dt} + \lambda_i\right)^{r-1} \cdot e_i(x_i)/i = 1, 2. \tag{10}$$

With:
λ_i : is a positive constant.
r : is the relative degree (Appendix 2).
And that for:

$$\begin{array}{lll} r = 1 & \Rightarrow & S(x) = e(x) \\ r = 2 & \Rightarrow & S(x) = \lambda_e(x) + \dot{e}(x) \\ r = 3 & \Rightarrow & S(x) = \lambda^2 e(x) + 2\lambda_{\dot{e}}(x) + \ddot{e}(x) \end{array} \tag{11}$$

The difference between the controlled variable and its reference is:

$$e_i(x) = x_i - x_{iref} \tag{12}$$

The purpose of this paper is to determine a control law to force the system states, i.e., the rotor flux and the electromagnetic torque to follow the sliding surface as: $S = [S_1 \quad S_2]^T$ (Ghanes and Zheng 2009).

4.2.2 Area Calculation

After the calculation of the relative degree (given in [Appendix 2]). The sliding surfaces of the Eq. (12), can be determined as follows:

$$\begin{cases} S_1 = \lambda_1 \cdot e_1 + \dot{e}_1 \\ S_2 = \lambda_2 \cdot e_2 + \dot{e}_2 \end{cases} \tag{13}$$

With :

$$\begin{cases} e_1 = \Phi_r - \Phi_{iref} \\ e_2 = \Omega_r - \Omega_{iref} \end{cases} \tag{14}$$

Are successively error flux (e_1) and error rate (e_2).

When substituting (1) and (14) into (13) the following result is as follows:

$$\begin{cases} S_1 = \lambda_1 \cdot (\Phi_r - \Phi_{iref}) + a_{31} \cdot (x_1 \cdot x_3 + x_2 \cdot x_4) + 2 \cdot a_{33} \cdot \Phi_r - \dot{\Phi}_{iref} \\ S_2 = \lambda_2 \cdot (\Omega_r - \Omega_{iref}) + \mu \cdot (x_2 \cdot x_3 - x_1 \cdot x_4) - \frac{c_r}{J} - \frac{f}{J}.\Omega_r - \dot{\Omega}_{iref} \end{cases} \tag{15}$$

However, to continue Φ_{iref} and Ω_{iref}, it suffices to make the sliding surface attractive and invariant.

4.2.3 Equivalent Command for the Invariance

Once the sliding surface is chosen, it remains to determine the control necessary to attract the controlled variable to the surface and then to his balance point, the relation will be as follows:

$$u = u_{eq} + u_n \tag{16}$$

where u_{eq} is called the equivalent control, which dictates the motion of the state trajectory along the sliding surface, and u_n is a term introduced to satisfy the condition following convergence $\dot{S}(x).S(x) < 0$, and it determines the dynamic behavior of the system during the convergence mode. So this command guarantees the attractiveness of the variable to be controlled to the sliding surface, (Utkin 1993; Dwards and Spurgeon 1998; Wai 2000).

The necessary condition for the system states follow the path defined by the sliding surfaces is:

$$\dot{S} = 0 \tag{17}$$

The equivalent command is the commands ensure the condition (17). Then the derivation of Eq. (15) gives:

$$\begin{cases} \dot{S}_1 = 2 \cdot (\lambda \cdot a_{33} + 2 \cdot a_{33}^2 + a_{31} \cdot a_{13}) \cdot \Phi_r + (\lambda \cdot a_{31} + 3 \cdot a_{33} \cdot a_{31} + a_{11} \cdot a_{31}) \cdot f_1 \\ \quad -(a_{31} \cdot a_{34}) \cdot f_2 + a_{31}^2 \cdot f_3 - \lambda \cdot \dot{\Phi}_{rref} - \ddot{\Phi}_{rref} + b_{11} \cdot a_{31} \cdot (x_3 \cdot U_1 \ + x_4 \cdot U_2\ ) \\ \dot{S}_2 = \mu \cdot \left(\lambda_1 + a_{33}^{+} a_{11}\right) \ \cdot f_2 - \lambda_2 . \frac{c_r}{J} + \cdot \mu \cdot a_{34} . x_5 \cdot f_1 + - (a_{34} \cdot a_{14} \cdot \mu) \cdot \Phi_r \\ \quad -\lambda_2 . \dot{\Omega}_{rref} - \ddot{\Omega}_{rref} + b_{11} \cdot \mu \cdot (x_3 \cdot U_1 \ - x_4 \cdot U_2\ ) \end{cases} \tag{18}$$

Such as:

$$\begin{cases} f_1 = x_1 \cdot x_3 + x_2 \cdot x_4 \\ f_2 = x_2 \cdot x_3 - x_1 \cdot x_4 \\ f_3 = x_1^2 + x_2^2 \end{cases} \tag{19}$$

The ideal diet is almost never possible. Therefore, the second term of the command must be used to restore the system state to the surface whenever it deviates. Thus, it should be taken as follows:

$$u_n = M_i \cdot sign\,(S_i(x)) \tag{20}$$

M_i is a constant, representing the maximum controller output required to overcome parameter uncertainties and disturbances; and $S_i\ (x)$ is called the switching function because the control action switches its sign on the two sides of the switching surface $S = 0$. A second-order system S is defined in Eq. (13), (Utkin 1993; Dwards and Spurgeon 1998; Wai 2000; Boucheta et al. 2012).

$S(x)$ Slip function is selected such that it is a solution of the following differential equation:

$$\dot{S}_i\ (x) = -M_i \cdot sign\ (S_i\ (x)) \ \ / \, i = 1, 2. \tag{21}$$

Then equation (17) can be written:

$$\begin{cases} -M_1 \cdot sign\,(S_1\,(x)) = 2 \cdot (\lambda \cdot a_{33} + 2 \cdot a_{33}^2 + a_{31} \cdot a_{13}) \cdot \Phi_r + (\lambda \cdot a_{31} + 3 \cdot a_{33} \cdot a_{31} + a_{11} \cdot a_{31}) \cdot f_1 \\ \qquad -(a_{31} \cdot a_{34}) \cdot f_2 + a_{31}^2 \cdot f_3 - \lambda \cdot \dot{\Phi}_{rref} - \ddot{\Phi}_{rref} + b_{11} \cdot a_{31} \cdot (x_3 \cdot U_1 \ + x_4 \cdot U_2\ ) \\ -M_2 \cdot sign\,(S_2\,(x)) = \mu \cdot (\lambda_1 + a_{33} + a_{11}) \ \cdot f_2 - \lambda_2 . \frac{c_r}{J} + \cdot \mu \cdot a_{34} . x_5 \cdot f_1 + - (a_{34} \cdot a_{14} \cdot \mu) \cdot \Phi_r \\ \qquad -\lambda_2 . \dot{\Omega}_{rref} - \ddot{\Omega}_{rref} + b_{11} \cdot \mu \cdot (x_3 \cdot U_1 \ - x_4 \cdot U_2\ ) \end{cases} \tag{22}$$

According to the Eqs. (19), (20) and (22) the equivalent command (16) for this invariance cans determined as (Araujo and Freitas 2000; Boucheta et al. 2012):

$$u = \begin{bmatrix} U_1 \\ U_2 \end{bmatrix} = G^{-1} \cdot \begin{bmatrix} X \\ Y \end{bmatrix} \tag{23}$$

With:

$$\begin{cases} X = 2\cdot(\lambda\cdot\left(\frac{a_{33}}{a_{31}}\right) + 2\cdot\left(\frac{a_{33}^2}{a_{31}}\right) + a_{13})\cdot\Phi_r + (\lambda + 3\cdot a_{33} + a_{11})\cdot f_1 - (a_{34})\cdot f_2 \\ \quad +a_{31}\cdot f_3 - \left(\frac{\lambda}{a_{31}}\right)\cdot\dot{\Phi}_{rref} - \left(\frac{1}{a_{31}}\right)\ddot{\Phi}_{rref} + \left(\frac{M_1}{a_{31}}\right)\cdot sign\,(S_1\,(x)) \\ Y = (\lambda_1 + a_{33} + a_{11})\ \cdot f_2 - \lambda_2.\frac{c_r}{J.\mu} + a_{34}.x_5\cdot f_1 + -(a_{34}\cdot a_{14})\cdot\Phi_r \\ \quad -\left(\frac{\lambda_2.}{\mu}\right).\dot{\Omega}_{rref} - \left(\frac{1}{\mu}\right)\ddot{\Omega}_{rref} + \left(\frac{M_2}{\mu}\right)\cdot sign\,(S_2\,(x)) \end{cases} \tag{24}$$

And

$$G = \begin{bmatrix} -b_{11}\cdot x_3 & -b_{11}\cdot x_4 \\ b_{11}\cdot x_4 & -b_{11}\cdot x_3 \end{bmatrix} \tag{25}$$

In taking into consideration the condition transversal matrix (25), then:

$$\det G \neq 0 \tag{26}$$

Therefore:

$$b_{11}^2\cdot\left(x_3^2 + x_4^2\right) \neq 0$$

With:

$$\begin{cases} x_3 = \phi_{r\alpha} \\ x_4 = \phi_{r\beta} \end{cases} \tag{27}$$

The determinant is non-zero, therefore, the matrix G is invertible, unless x = 0 and/or stopping the motor; the current is zero so the flux is zero, there must be provided the initial conditions of flux at startup.

For the switching law to intervene in the law of overall control, choose and sufficiently large, convergence criteria (Aurora and Ferrara 2007; Boucheta et al. 2012).

$$\begin{cases} M_1 > \left| \begin{matrix} 2\cdot\left(\lambda\cdot\left(\frac{a_{33}}{a_{31}}\right) + 2\cdot\left(\frac{a_{33}^2}{a_{31}}\right) + a_{13}\right)\cdot\Phi_r \\ +(\lambda + 3\cdot a_{33} + a_{11})\cdot f_1 \\ -(a_{34})\cdot f_2 + a_{31}^{\cdot} f_3 - \left(\frac{\lambda}{a_{31}}\right)\cdot\dot{\Phi}_{rref} \\ -\left(\frac{1}{a_{31}}\right)\ddot{\Phi}_{rref} \end{matrix} \right| \\ M_2 > \left| \begin{matrix} (\lambda_1 + a_{33} + a_{11})\ \cdot f_2 - \lambda_2.\frac{c_r}{J.\mu} \\ +a_{34}.x_5\cdot f_1 - (a_{34}\cdot a_{14})\cdot\Phi_r \\ -\left(\frac{\lambda_2.}{\mu}\right).\dot{\Omega}_{rref} - \left(\frac{1}{\mu}\right)\ddot{\Omega}_{rref} \end{matrix} \right| \end{cases} \tag{28}$$

The diagram of sliding mode control is given in Fig. 5 (Saiad 2012):

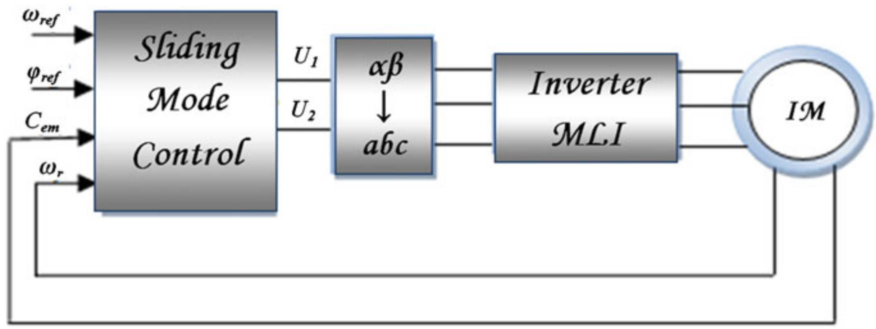

Fig. 5 Block diagram of the sliding mode control

5 Simulations Results

The simulation analysis of the mathematical model of the induction machine, were carried out in MATLAB/SIMULINK to demonstrate the effectiveness of the proposed control scheme for speed control of the IM described above, and allows seeing the performance comparison of control: Sliding Mode Control (SMC) and field oriented control (FOC), with test of two mode of operation, speed variation and other inversion of rotation with variable load torque C_r and rotor resistance R_r.

5.1 Simulation Results Without Inversion Speed

This simulation is realized with the reference speed given in Table 1 as follow:

5.1.1 Analysis and Discussions of the Results

The results simulations are given in the following Figs. 6, 7, 8, 9, 10, 11 12:

The proposed controller has been tested also with detuned rotor resistance. The rotor resistance is considered the most effective parameter in the indirect vector control as the slip calculator depends mainly on it. In classical indirect vector control, the variation of this parameter will adversely affect the motor performance. In this test, the rotor resistance is assumed to increase as follows: at 0.8 s will rise to 50 %

Table 1 The reference speed at the first simulation

Times t (s)	$0 \rightarrow 1.5$	$1.5 \rightarrow 2.75$	$2.75 \rightarrow 4$
Reference speed Ω_{ref} (rad/s)	157	170	100

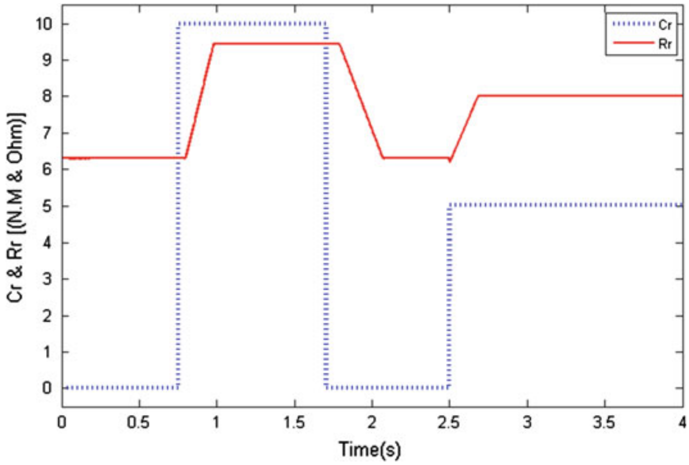

Fig. 6 Load and rotor resistance variations

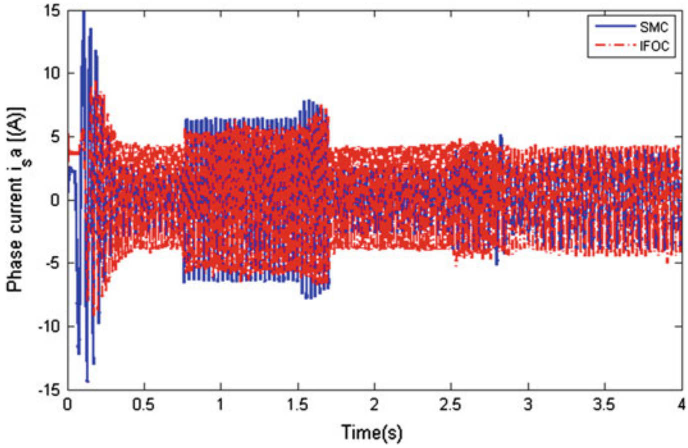

Fig. 7 Current of one phase

R_r , and at 2.1 s will increase to 30 % R_r (Fig. 6), in the machine model and keeping nominal value in the slip calculator.

In this test too, the load torque is assumed to change from 0 to 10 Nm. at 0.75 s and stepped again to no load at 1.75 s, and change again from 0 to 5 Nm. at 2.5 s as shown in Fig. 6.

The Fig. 7 reports an enlarged view of the phase stator current during high speed operation. It is seen that, sinusoidal current waveform is obtained with less distortion.

The Fig. 8 indicates that the control Sliding Mode Control (SMC) provides a successful prosecution at its rotor flux reference (unlike the control Indirect Field Oriented Control (IFOC)). Thus, the d-axis rotor flux linkage is kept constant at

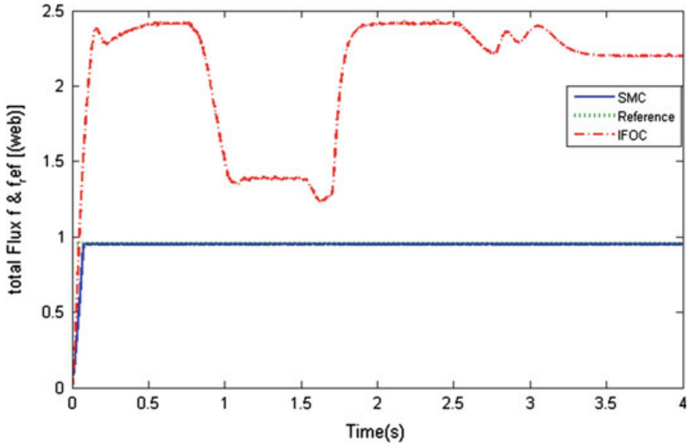

Fig. 8 Rotor flux tracking performance

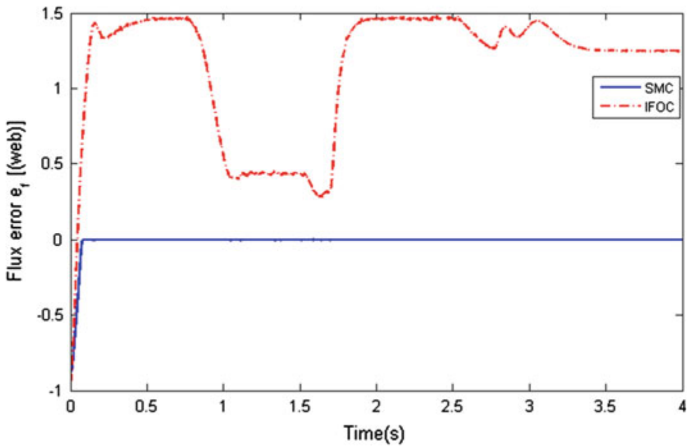

Fig. 9 Rotor flux error

the rated value, while the q-axis flux is kept zero in all the simulation period; only, small notches in the d-axis flux have been noticed at the instants of load disturbance application, (in the static regime the error is zero as shown in Fig. 9). In other words, the decoupling condition between the speed and rotor flux has been realized.

Figure 10 shows the results of simulation speed with the two types of controls (IFOC and SMC), when the machine is operated at different speeds as it's given in Table 1. It can be seen that the speed follows its reference reasonably well, but the SMC controller has a response time better than IFOC (0.35s), this time can be explained by the speed of this technique. As shown in Fig. 10 Zoom 1.

By applying a resisting torque which produces the heating of the machine, who led to varied the rotor see Fig. 6. In these circumstances, finds that the speed perfectly

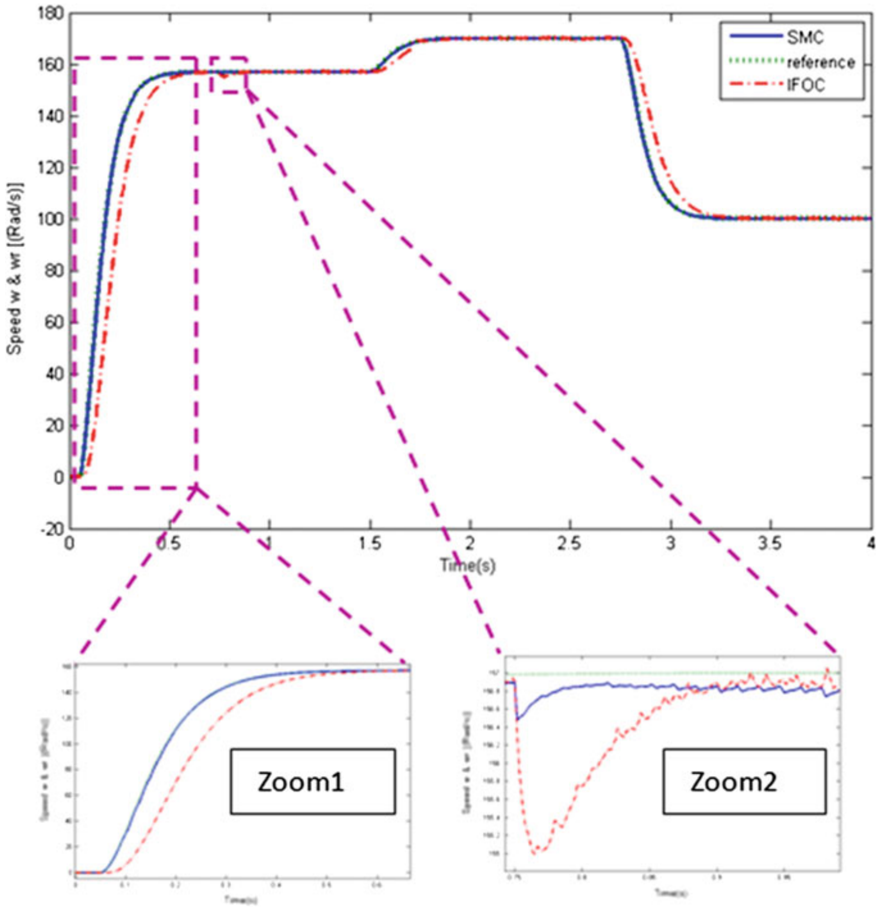

Fig. 10 Rotor speed tracking performance

follows its reference in the two types of control that is clearly demonstrated by the speed error is zero see Fig. 11, but here too the SMC has a time response is perfect and the maximum overshoot is about 2 % compared to IFOC as shown in Fig. 10 Zoom 2, good tracking performance has been achieved with the proposed sliding controller in spite of the mismatched rotor resistance.

It has been indicated in the Fig. 12 that excellent tracking performance has been achieved in spite of the load torque disturbance. Only, small notches have been noticed at the instants of load disturbance application, peaks 30 % of the load torque for SMC and 20 % for IFOC.

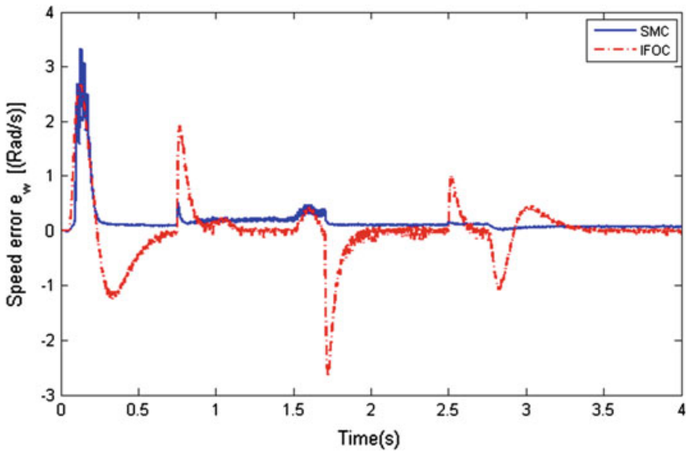

Fig. 11 Rotor speed error

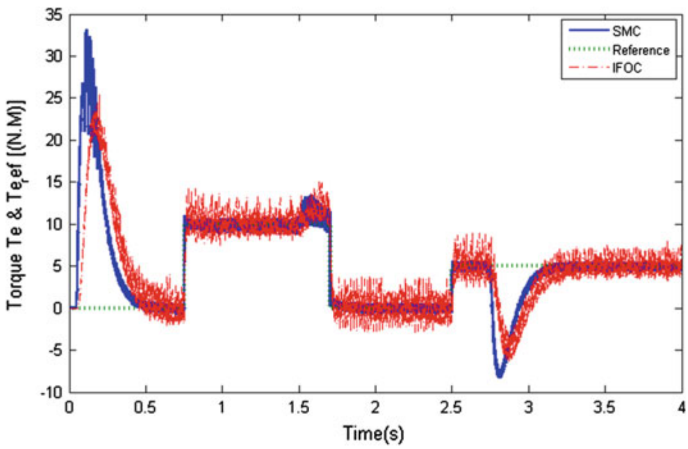

Fig. 12 Electromechanical torque to the load variation

5.2 Simulation Results with Inversion Speed

This simulation is realized with the reference speed given in Table 2 as follow:

5.2.1 Analysis and Discussions of the Results

The results simulations are given as following Figs. 13, 14, 15, 16, 17, 18 19:

The proposed controller has been tested also with detuned rotor resistance. The rotor resistance is considered the most effective parameter in the indirect vector

Table 2 The reference speed variation at the second simulation

Times t (s)	$0 \rightarrow 1.5$	$1.5 \rightarrow 3$	$3 \rightarrow 4$
Reference speed Ω_{ref} (rad/s)	157	−157	157

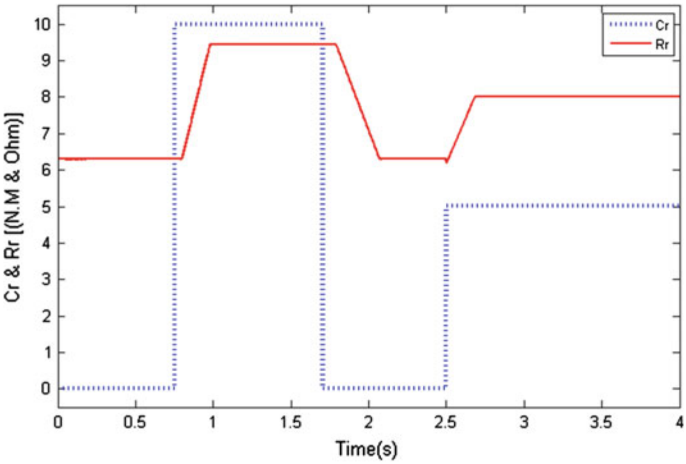

Fig. 13 Load and rotor resistance variations

control as the slip calculator depends mainly on it. In classical indirect vector control, the variation of this parameter will adversely affect the motor performance. In this test, the rotor resistance is assumed to increase as follows: at 0.8 s will rise to 50 % R_r , and at 2.1 s will increase to 30 % R_r (Fig. 13), in the machine model and keeping nominal value in the slip calculator.

In this test too, the load torque is assumed to change from 0 to 10 Nm. at 0.75 s and stepped again to no load at 1.75 s, and change again from 0 to 5 Nm. at 2.5 s as shown in Fig. 13.

The Fig. 14 reports an enlarged view of the phase stator current during high speed operation. It is seen that, sinusoidal current waveform is obtained with less distortion for both techniques, with peaks in the transitional regime and at the inversion of speed.

The Fig. 15 indicates that the control Sliding Mode Control (SMC) provides a successful prosecution at its rotor flux reference (unlike the control Indirect Field Oriented Control (IFOC)). Thus, the d-axis rotor flux linkage is kept constant at the rated value, while the q-axis flux is kept zero in all the simulation period; only, small notches in the d-axis flux have been noticed at the instants of load disturbance application and moment of inversion the direction of rotation, (in the static regime the error is zero as shown in Fig. 16). In other words, the decoupling condition between the speed and rotor flux has been realized.

Figure 17 shows the results of simulation speed with the two types of controls (IFOC and SMC), when the machine is operated at different speeds as it's given in Table 2. It can be seen that the speed follows its reference reasonably well, but

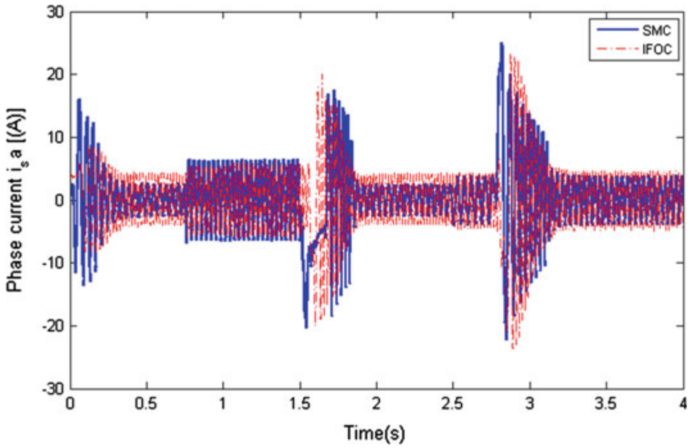

Fig. 14 Current of one phase

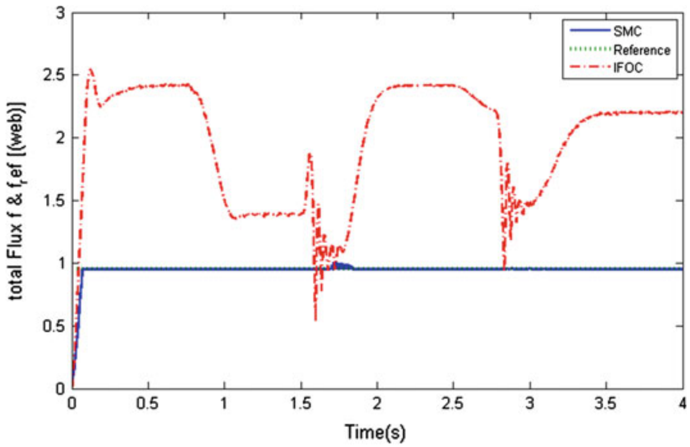

Fig. 15 Rotor flux tracking performance

the SMC controller has a response time better than IFOC (0.35s), this time can be explained by the speed of this technique. As shown in Fig. 17 Zoom 1.

By applying a resisting torque which produces the heating of the machine, who led to varied the rotor see Fig. 13. In those circumstances, finds that the speed perfectly follows its reference in the two types of control that is clearly demonstrated by the speed error is zero see Fig. 18, but here too the SMC has a time response is perfect and the maximum overshoot is about 2 % compared to IFOC as shown in Fig. 17 Zoom 2, good tracking performance has been achieved with the proposed sliding controller in spite of the mismatched rotor resistance and the reversing the direction of rotation.

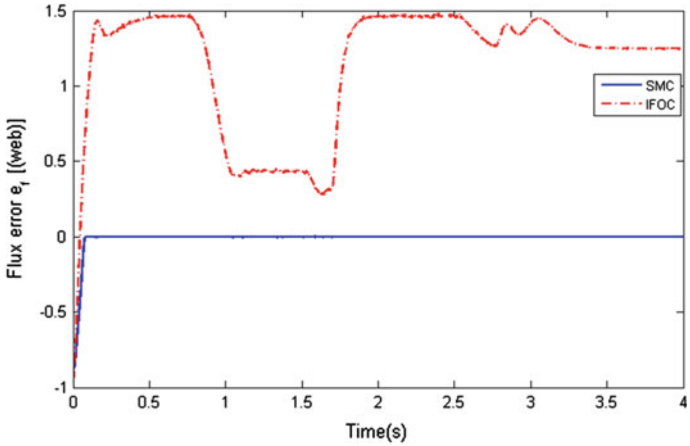

Fig. 16 Rotor flux error

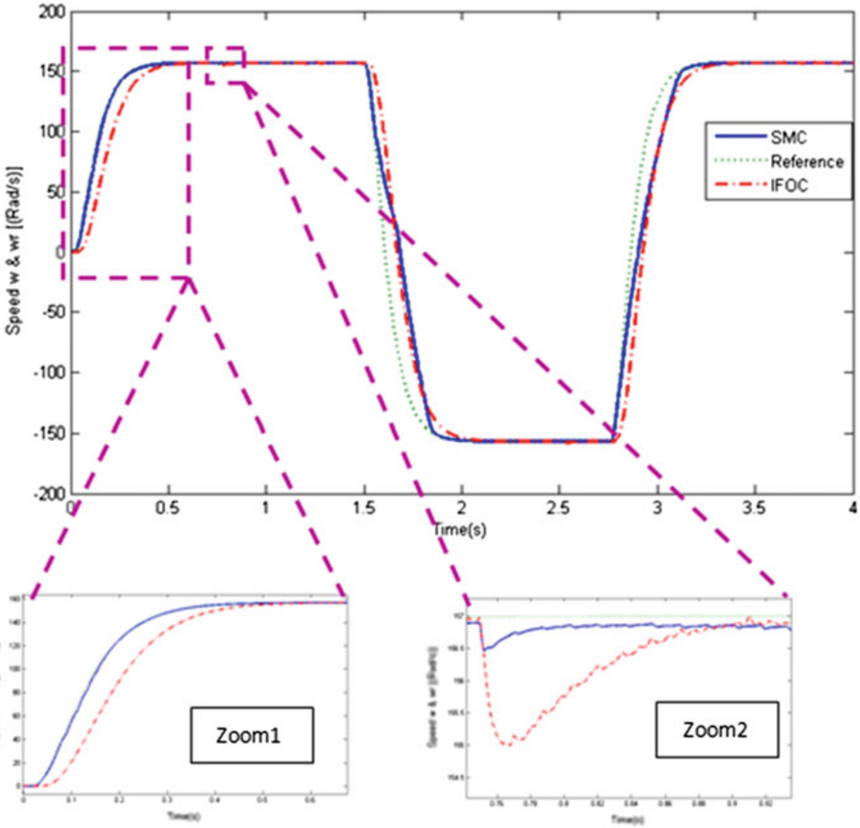

Fig. 17 Rotor speed tracking performance

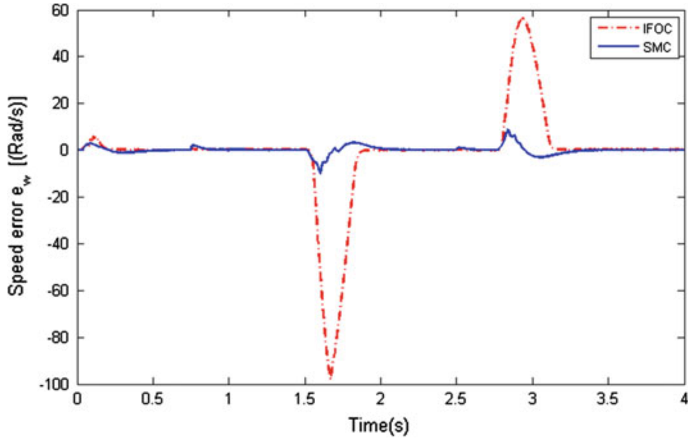

Fig. 18 Rotor speed error

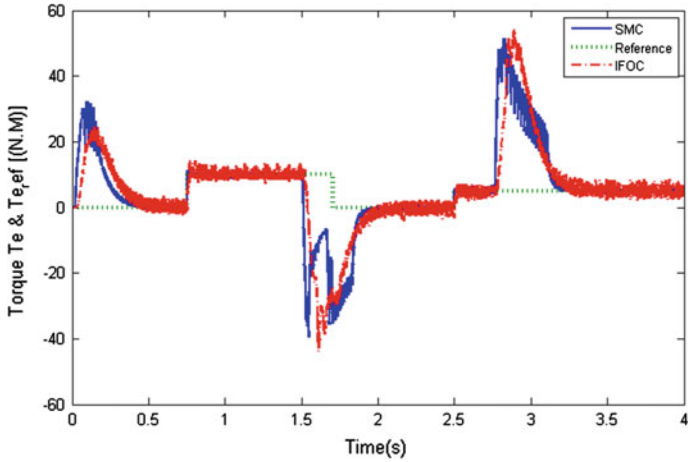

Fig. 19 Electromechanical torque to the load variation

It has been indicated in the Fig. 12 that excellent tracking performance has been achieved in spite of the load torque disturbance in spite of the mismatched rotor resistance and the reversing the direction of rotation. Only, small notches have been noticed at the instants of load disturbance application, peaks 30 % of the load torque for SMC and 20 % for IFOC.

6 Conclusion

Industrial systems are often significantly nonlinear behavior. The linearization around an operating point is often inadequate for the needs of the command, therefore it is important to develop control methods for nonlinear systems.

Controlling an IM can be done using several techniques, each of which offers dynamic and static performances with well-defined limits applications. The problem arises in the choice of a particular method. The use of a method or the other is normally based on the constraints of the specifications, which are sometimes added new requirements of energy saving and material economy that should be taken into account.

It is with this understanding that this work has been made. Indeed, the main objective of this chapter is the development of a new robust control by sliding mode. This type of control has been sufficiently discussed compared to the vector control of rotor flux orientation.

First vector control by indirect rotor flux orientation gave lower performance. Indeed, the strength of the test IFOC which has opposite variation of the rotor resistance shows that control loses its linearity property and affects more decoupling between rotor flux and torque. To improve the performance of this command and achieve better results, the online identification of parameters of the machine is essential.

In this sense, our contribution is to propose a methodology of robust control systems related to variable structures whose purpose is to overcome the disadvantages of conventional controls, as the SMC is by nature a non-linear control and their control law is changed in a discontinuous manner, in this case, the sliding mode control. This control is characterized by its robustness against external and internal disturbances. The sliding surface is determined depending on the desired performance. While, the control law is chosen in order to ensure the convergence conditions and sliding i.e., attractiveness and the invariance of switching surfaces.

Finally, the study of the sliding mode control of the induction machine consists in defining a sliding surface on which the system converges. The corresponding switching function allows the system to always tend towards the sliding surface. The technique of sliding mode control used for control of the induction motor has led to good performance, in many cases obtained in a better quality of adjustment relative to the vector control, it offers some advantages: first, robustness with respect to variations of system parameters, second, a high-performance dynamic "acceptable response time and error stationary practically zero", and finally a simple implementation of the law switching.

In addition, the objective of this work is achieved, because there is a perfect prosecution of the rotor flux, for its reference which makes for a good decoupling between the rotor flux and the electromagnetic torque, good trajectory tracking the desired output speed with a minimum response time, the robustness to variations in parameters, which has been shown by simulation results. The performance of this technique depends on a suitable choice of the coefficients of the sliding surface and the speed of the response depends on the maximum torque that can give the machine.

Perspectives From this, labor perspective can be considered, it would be interesting to:

- Propose a synthesis of an observer flux sliding mode. Because in the design of the sliding mode control, it is assumed that all states were measured. Since only the current measurements are available, then the estimated for the rotor flux of an application in real time is necessary.
- In an effort to reduce cost, speed sensor can be replaced by a flux observer and speed.
- And finally, it would be interesting to use this machine (induction machine IM) and this type of control (Sliding Mode Control SMC) for the adopted to renewable energies (wind).

Appendix: A.1 Arbitrary Reference Frame Theory

Arbitrary reference frame theory is mainly used in the dynamic analysis of electrical machines. Because of the highly coupled nature of the machine, especially the inductances within the winding make it rather impossible to perform dynamic simulations and analysis on electrical machines.

Arbitrary reference frame theory was discovered by Blondel, Dreyfus, Doherty and Nickle as mentioned in the classical paper (Park 1929). This newly found theory was generalized by Park on synchronous machines and this method was later extended by Stanley to the application of dynamic analysis of induction machines (Stanley 1938).

By using this method a poly-phase machine is transformed to a two-phase machine with their magnetic axis in quadrature as illustrated in Fig. 20. This method is also commonly referred to as the dq method in balanced systems and to the dq0 method in unbalanced systems with the '0' relating to the zero sequence or homopolar components in the Fortes cue Transformation (Jimoh et al. 2012).

This transformation eliminates mutual magnetic coupling between the phases and therefore makes the magnetic flux linkage of one winding independent of the current of another winding.

The transformation is done by applying a transformation matrix, Eq. (29) while the inverse transformation matrix, Eq. (30) will be transformed back to the natural reference frame. Eqs. (29) and (30) applly to a three phase system but can be modified to accommodate a system with any number of phases which might be useful in the case of the machine having an auxiliary winding as proposed in this work (Jimoh et al. 2012).

$$[P] = \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ \cos\theta & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos\left(\theta + \frac{2\pi}{3}\right) \\ -\sin\theta & -\sin\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) \end{bmatrix} \tag{29}$$

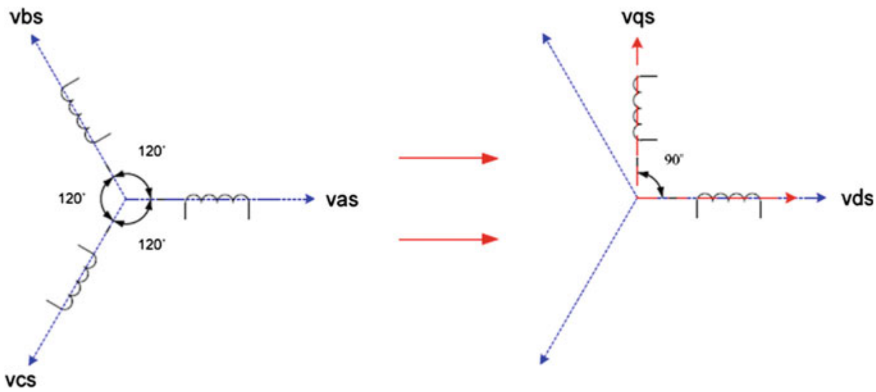

Fig. 20 Park's transform

$$[P]^{-1} = \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & \cos\theta & -\sin\theta \\ \frac{1}{\sqrt{2}} & \cos\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta - \frac{2\pi}{3}\right) \\ \frac{1}{\sqrt{2}} & \cos\left(\theta + \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) \end{bmatrix} \tag{30}$$

A.2 Modeling of Three-Phase Induction Motor

The winding arrangement of a symmetrical induction machine is shown in Fig. 21. The stator windings are identical and sinusoidally distributed, displaced 120° apart, with *Ns* equivalent turns and resistance *Rs* per winding, per phase. Similarly the rotor windings are also considered as three identical sinusoidally distributed windings, displaced 120° apart, with *Nr* equivalent turns and resistance of *Rr* per, winding per phase (Jimoh et al. 2012).

In developing the equations which describe the behaviour of the induction machine the following assumptions are made (Jimoh et al. 2012):

1. The airgap is uniform.
2. Eddy currents, friction and windage losses and saturation are neglected.
3. The windings are distributed sinusoidally around the air gap.
4. The windings are identical

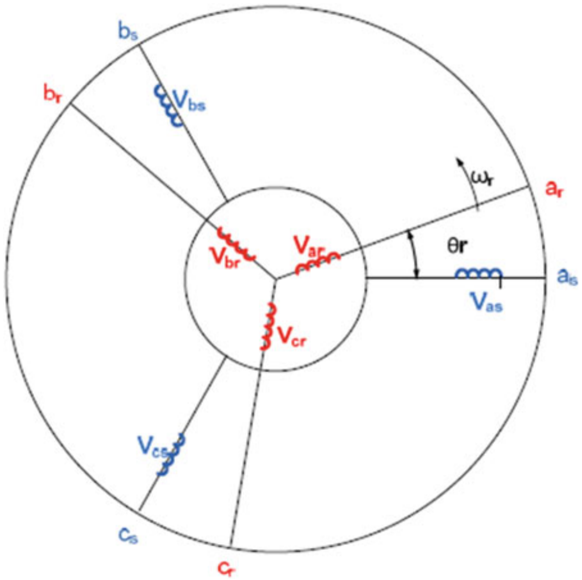

Fig. 21 Three-phase winding Arrangement

Appendix: B Calculates of the Relative degree

B.1 Derivation of Lie

Is $h : \Re^n \to \Re$ a scalar function and $f = \begin{bmatrix} f_1 & f_2 & \ldots & f_n \end{bmatrix}^T : \Re^n \to \Re^n$ a vector field with n is the order of the system (Pietrzak-David et al. 2000; Meziane et al. 2008).

We use the notation $L_f \cdot h\,(x) : \Re^n \to \Re$ to denote the given scalar function:

$$L_f \cdot h\,(x) = \frac{\partial h}{\partial x} \cdot f\,(x) = \left[\frac{\partial h}{\partial x_1}, \ldots, \frac{\partial h}{\partial x_n}\right] \cdot \begin{bmatrix} f_1\,(x) \\ \vdots \\ f_n\,(x) \end{bmatrix} = \sum_{i=1}^{n} \frac{\partial h}{\partial x_i} \cdot f_i\,(x) \quad (31)$$

where: $x = [x_1, x_2, \ldots, x_n]^T$

And $L_f \cdot h\,(x)$ is called the Lie derivative in the direction of the vector field f.

Similarly, it may be noted, for $k = 0, 1, 2, 3 \ldots$:

$$\begin{cases} L_f^k \cdot h\,(x) = \frac{\partial^k h}{\partial x^k} \cdot f\,(x) = \frac{\partial}{\partial x}\left(L_f^{k-1} \cdot h(x)\right) \cdot f\,(x) \\ L_f^0 \cdot h\,(x) = h(x) \end{cases} \quad (32)$$

B.2 Relative Degree

Consider the following nonlinear system:

$$\begin{cases} \overset{\bullet}{x} = f(x) + g(x)u \\ y = h(x) \end{cases} \tag{33}$$

Now consider the output $y = h\,(x)^{\in} \Re$. It is said that $y = h\ (x)$ has a degree relative r with respect to the input scalar u where:

$$\begin{cases} L_g L_f^k \cdot h\,(x) = 0\ ,\ \ 0 \leq k \leq r - 1 \\ L_g L_f^{r-1} \cdot h\,(x) \neq 0 \end{cases} \tag{34}$$

Lie derivative of the scalar function $h\,(x)$ taken along first and then along the second vector g is defined by:

$$L_g L^{\cdot}_f h\,(x) = \frac{\partial\left(L_f h\right)}{\partial x} \cdot g\,(x) \tag{35}$$

Notes

- The concept of relative degree r is very important during the linearization because it lets us know if our system is linearizable completely or partially.
- It should be noted that the relative degree r is the number of times to derive the output y for the u control appears, it is verified as follows:

$$\dot{y} = \frac{\partial h}{\partial x}\dot{x} = \frac{\partial h}{\partial x} \cdot f\,(x) + \frac{\partial h}{\partial x} \cdot g(x) \cdot u \tag{36}$$

$$\dot{y} = L_f \cdot h\,(x) + L_g \cdot h\,(x) \cdot u \tag{37}$$

And since $L_g \cdot h\,(x) = 0$, then:

$$\dot{y} = L_f \cdot h\,(x) \tag{38}$$

Likewise we find:

$$\begin{aligned} &\ddot{y} = L_f^2 \cdot h\,(x) \\ &\vdots \\ &y^{(r-1)} = L_f^{(r-1)} \cdot h\,(x) \\ &y^r = \tfrac{\partial}{\partial x}\left(L_f^{(r-1)} \cdot h\,(x)\right)\overset{\bullet}{x} \\ &\ \ = L_f^r \cdot h\,(x) + L_g \cdot L_f^{r-1} \cdot h\,(x) \cdot u \end{aligned} \tag{39}$$

Calculate the Relative Degree of the Induction Motor System The relative degree r_i of system corresponding to outputs y_i of the induction machine to the relationship (33) is given as follows:

(a) *Relative degree of module rotor flux*

$$h_1(x) = (x_3^2 + x_4^2) \tag{40}$$

From Eq. (36), derived the Eq. (40) found:

$$\dot{h}_1(x) = a_{31}.(x_1.x_3 + x_2x_4) + a_{33}.(x_3^2 + x_4^2) \tag{41}$$

The first derivative of h_1 *(x)* does not involve the input v. then it must derive a second time, as find in Eq. (39).

$$\dot{h}_1 = (x_1x_3 + x_2x_4).(a_{11}a_{33} + 3a_{33}a_{31}) - (x_2x_3 - x_1x_4).(a_{31}a_{34}) + (x_3^2 + x_4^2). \\ (2.a_{33}^2 + a_{13}a_{31}) + a_{31}b_{11}x_3U_1 + a_{31}b_{11}x_4U_2 \tag{42}$$

The U_1, U_2 commands appear after the second derivative; therefore, the relative degree with respect to $h_1(x)$ is $r_1 = 2$.

(b) *Relative degree of the rotation speed.*

$$h_2(x) = \Omega_r \tag{43}$$

From Eq. (36), derived the Eq. (43) found:

$$\dot{h}_2(x) = \mu.(x_2.x_3 - x_1.x_4) - \frac{c_r}{J} \tag{44}$$

Again the first derivative of h_2 *(x)* does not involve the input v. then it must derive a second time, as find in Eq. (39).

$$\ddot{h}_2 = \dot{\Omega} = J \cdot \mu \cdot a_{34} \cdot (x_1 \cdot x_3 + x_2 \cdot x_4) - J \cdot \mu \cdot a_{14} \cdot \left(x_3^2 + x_4^2\right) + +J \cdot \mu \cdot b_{11} \\ (x_3 \cdot U_2 - x_4 \cdot U_1) + J \cdot \mu \cdot (a_{11} + a_{33}) \cdot (x_2 \cdot x_3 - x_1 \cdot x_4) \tag{45}$$

The U_1, U_2 commands appear at the end of the second derivative, so the relative degree with respect to h_2 (x) is $r_2 = 2$.

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