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Cascade Sliding Mode Control of a Field Oriented Induction Motors with Varying Parameters

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1. Introduction

The induction machine is widely used in industry, because of its mechanical robustness, low maintenance requirement, and relatively low cost. However, from control point of view, control of the induction machines is one of the most challenging topics. Its control is complex because the dynamic of the induction machine is nonlinear, multivariable, and highly coupled. Furthermore, there are various parameter uncertainties and disturbances in the system. The rotor resistances, for example, can vary up to 100% because of rotor heating during operation. In the last few years, many versions of a nonlinear state feedback control schemes, such as, input-output feedback linearization ((Marino et al., 1993)), passivity-based control ((Ortega & Espinoza, 1993; Ortega et al., 1996)) and Backstepping (Kanellakopoulos et al. (1991); Krstic et al. (1995)) have been applied to the IM drive. Adaptive versions of most of those nonlinear control schemes are also available for the effective compensation of the parameter uncertainties and disturbances in the induction motor systems (Ebrahim & Murphy (2006); Marino et al. (1993); Ortega et al. (1993); Rashed et al. (2006)). A fundamental problem in the design of feedback controllers is that of stabilizing and achieving a specified transient performance in the presence of external disturbances and plant parameter variations.

Since the publication of the survey paper by (Utkin (1977)), significant interest on Sliding mode control has been generated in the control research community worldwide. This interest is increased in the last two decades due to the possibility to implement this control in industrial applications with the advances of the power electronics technology and the availability of cheap and fast computation.

One of the most intriguing aspects of sliding mode is the discontinuous nature of the control action whose primary function of each of the feedback channels is to switch between two distinctively different system structures (or components) such that a new type of system motion, called sliding mode, exists in a manifold. This peculiar system characteristic is claimed to result in superior system performance which includes insensitivity to parameter variations, and complete rejection of disturbances (Young et al. (1999)).

In this paper, a nonlinear adaptive Sliding mode speed and rotor flux control scheme combined with field orientation for the induction-motor drive has been developed. Some sliding surfaces are chosen for which an appropriate logic commutation associated to these surfaces is determined. One important characteristic of the proposed controller is its cascade structure; which gives a high performance using simple sliding surfaces. Furthermore, in order to reduce chattering phenomenon, smooth control functions with appropriate threshold have

been chosen. The rotor flux is estimated using the rotor-circuit model and, thus, is insensitive to the stator resistance. A stable model reference adaptive system (MRAS) rotor-resistance and load torque estimators have been designed using the measured stator current and rotor speed, and voltage command.

2. Indirect field-oriented control of the IM

Assuming linear magnetic circuits and balanced three phase windings, the fifth-order nonlinear model of IM (Krauss (1995)), expressed in the stator frame is :

$$\begin{aligned}
 \frac{d\omega}{dt} &= \frac{3n_p L_m}{2JL_2} (\lambda_{2a} i_{1b} - \lambda_{2b} i_{1a}) - \frac{T_L}{J} \\
 \frac{d\lambda_{2a}}{dt} &= -\frac{R_2}{L_2} \lambda_{2a} - n_p \omega \lambda_{2b} + \frac{R_2}{L_2} L_m i_{1a} \\
 \frac{d\lambda_{2b}}{dt} &= -\frac{R_2}{L_2} \lambda_{2b} + n_p \omega \lambda_{2a} + \frac{R_2}{L_2} L_m i_{1b} \\
 \frac{di_{1a}}{dt} &= \frac{L_m R_2}{\sigma L_1 L_2^2} \lambda_{2a} + \frac{n_p L_m}{\sigma L_1 L_2} \omega \lambda_{2b} - \frac{L_m^2 R_2 + L_2^2 R_1}{\sigma L_1 L_2^2} i_{1a} + \frac{1}{\sigma L_1} u_{1a} \\
 \frac{di_{1b}}{dt} &= \frac{L_m R_2}{\sigma L_1 L_2^2} \lambda_{2b} - \frac{n_p L_m}{\sigma L_1 L_2} \omega \lambda_{2a} - \frac{L_m^2 R_2 + L_2^2 R_1}{\sigma L_1 L_2^2} i_{1b} + \frac{1}{\sigma L_1} u_{1b}
 \end{aligned} \tag{1}$$

We can see that the model described by (1) is highly coupled, multivariable and nonlinear system. It is very difficult to control the IM directly based on this model. State transformation to simplify the system representation is required. A well-known method to this end is the transformation of the field orientation principle. It involves basically a change of the representations of the state vector $(i_{1a}, i_{1b}, \lambda_{2a}, \lambda_{2b})$ in the fixed stator frame (a, b) into a new state vector in a frame (d, q) which rotates along with the flux vector $(\lambda_{2a}, \lambda_{2b})$. Mathematically, the field oriented transformation can be described as:

$$\begin{aligned}
 i_{1d} &= \frac{\lambda_{2a} i_{1a} + \lambda_{2b} i_{1b}}{\sqrt{\lambda_{2a}^2 + \lambda_{2b}^2}}, & i_{1q} &= \frac{\lambda_{2a} i_{1b} - \lambda_{2b} i_{1a}}{\sqrt{\lambda_{2a}^2 + \lambda_{2b}^2}} \\
 \lambda_{2d} &= \sqrt{\lambda_{2a}^2 + \lambda_{2b}^2}, & \lambda_{2q} &= 0, & \rho &= \arctan\left(\frac{\lambda_{2b}}{\lambda_{2a}}\right)
 \end{aligned} \tag{2}$$

Where much simplification is gained by the fact that $\lambda_{2q} = 0$. Using this transformation and the notations in the Nomenclature, the state equations (1) can be rewritten in the new state variables as:

$$\begin{aligned}
 \frac{d\omega}{dt} &= \mu \lambda_{2d} i_{1q} - \frac{T_L}{J} \\
 \frac{di_{1q}}{dt} &= -\eta_1 i_{1q} - \beta n_p \omega \lambda_{2d} - n_p \omega i_{1d} - R_2 (\eta_2 i_{1q} + \alpha L_m \frac{i_{1q} i_{1d}}{\lambda_{2d}}) + \frac{1}{\sigma L_1} u_{1q} \\
 \frac{d\lambda_{2d}}{dt} &= -\alpha R_2 \lambda_{2d} + \alpha L_m R_2 i_{1d}
 \end{aligned}$$

$$\frac{di_{1d}}{dt} = -\eta_1 i_{1d} + n_p \omega i_{1q} + R_2(-\eta_2 i_{1d} + \alpha \beta \lambda_{2d} + \alpha L_m \frac{i_{1q}^2}{\lambda_{2d}}) + \frac{1}{\sigma L_1} u_{1d} \tag{3}$$

$$\frac{d\rho}{dt} = n_p \omega + \alpha L_m R_2 \frac{i_{1q}}{\lambda_{2d}}$$

The decoupling control method with compensation is to choose inverter output voltages such that:

$$u_{1d}^* = \left(K_p + K_i \frac{1}{s} \right) (i_{1d}^* - i_{1d}) - \rho L_1 \sigma i_{1q}^* + L_1 \sigma \frac{di_{1q}^*}{dt}$$

$$u_{1q}^* = \left(K_p + K_i \frac{1}{s} \right) (i_{1q}^* - i_{1q}) + \rho L_1 \sigma i_{1d}^* + \rho \frac{L_m}{L_2} \hat{\lambda}_{2d} \tag{4}$$

Where K_p, K_i are PI controller gains.

For that, we need the estimation of the rotor flux as given by

$$\hat{\lambda}_{2d} = -\alpha \hat{R}_2 \hat{\lambda}_{2d} + \alpha L_m \hat{R}_2 i_{1d} \tag{5}$$

Fig.1 shows the implemented diagram of an induction motor indirect field-oriented control (IFOC).

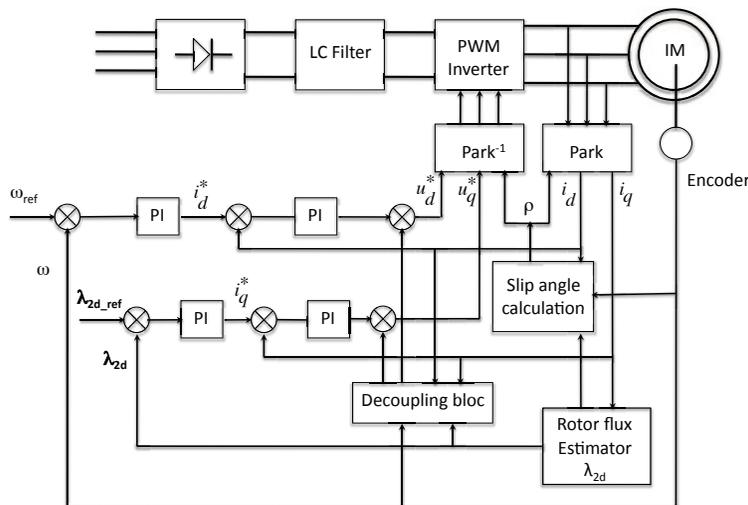


Fig. 1. bloc diagram of IFOC for the IM

3. Cascade sliding mode control

From the above IFOC structure, the sliding mode control mechanisms for the rotor angular speed regulation and the flux generation can be better applied to replace the traditional nonlinear feedback PI control of the field oriented control technique for better performance. Basically, the sliding mode control objectives consist mainly of the following steps:

3.1 Design of the switching surfaces

We choose the sliding surface to obtain a sliding mode regime which guarantees the convergence of the state x to its desired value x_d according to the relation (6) (Slotine & Li (1991)):

$$S(x) = \left(\frac{d}{dt} + \lambda \right)^{r-1} e \quad (6)$$

Where $e = x_d - x$: tracking error

λ : positive coefficient

r : relative degree

Such in IFOC structure, we have four PI controllers; we will define four sliding surfaces (Mahmoudi et al. (1999)):

$$\begin{aligned} S_1(\omega) &= \omega_{ref} - \omega \\ S_2(\lambda_{2d}) &= \lambda_{2dref} - \lambda_{2d} \\ S_3(i_{1q}) &= i_{1q}^* - i_{1q} \\ S_4(i_{1d}) &= i_{1d}^* - i_{1d} \end{aligned} \quad (7)$$

3.2 Control calculation

Two parts have to be distinguished in the control design procedure. The first one concerns the attractivity of the state trajectory to the sliding surface and the second represents the dynamic response of the representative point in sliding mode. This latter is very important in terms of application of nonlinear control techniques, because it eliminates the uncertain effect of the model and external perturbation. For that, the structure of a sliding mode controller includes two terms:

$$u_c = u_{eq} + u_n \quad (8)$$

Where

$-u_{eq}$ is called equivalent control which is used when the system state is in the sliding mode. It is calculated from $\dot{S}(x) = 0$.

$-u_n$ is given to guarantee the attractivity of the variable to be controlled towards the commutation surface. This latter is achieved by the condition (Slotine & Li (1991); Utkin (1993)).

$$S(x) \cdot \dot{S}(x) < 0 \quad (9)$$

A simple form of the control action using sliding mode theory is a relay function; which has a discontinuous form given by:

$$u_n = -k \cdot \text{sgn}(S(x)) \quad (10)$$

k is a constant and is chosen positive to satisfy attractivity and stability conditions.

3.2.1 For the rotor flux regulation

$$\dot{S}_2(\lambda_{2d}) = 0 \Rightarrow i_{1deq} = \frac{\lambda_{2d} + T_r \dot{\lambda}_{2dref}}{L_m} \quad (11)$$

$$\dot{S}_2(\lambda_{2d}) S_2(\lambda_{2d}) < 0 \Rightarrow i_{1dn} = K_\phi \cdot \text{sign}(S_2(\lambda_{2d})) \quad (12)$$

Thus the controller is

$$i_{1dc} = i_{1deq} + i_{1dn} \quad (13)$$

3.2.2 For the direct current regulation

$$\begin{aligned} \dot{S}_4(i_{1d}) = 0 \Rightarrow \\ u_{1deq} = \sigma L_1 \dot{i}_{1dref} + R_{sm} i_{1d} - \sigma L_1 \omega_s i_{1q} - \frac{L_m}{L_2 T_r} \lambda_{2d} \end{aligned} \quad (14)$$

$$\dot{S}_4(i_{1d}) S_4(i_{1d}) < 0 \Rightarrow u_{1dn} = K_d \cdot \text{sign}(S_4(i_{1d})) \quad (15)$$

The controller is given by

$$u_{1dc} = u_{1deq} + u_{1dn} \quad (16)$$

3.2.3 For the speed regulation

$$\dot{S}_1(\omega) = 0 \Rightarrow i_{1qeq} = \frac{J \dot{\omega}_{ref} + F \omega + T_L}{\frac{p L_m}{L_2} \lambda_{2d}} \quad (17)$$

$$\dot{S}_1(\omega) S_1(\omega) < 0 \Rightarrow i_{1qn} = K_\omega \cdot \text{sign}(S_1(\omega)) \quad (18)$$

The controller is given by

$$i_{1qc} = i_{1qeq} + i_{1qn} \quad (19)$$

3.2.4 For the quadrature current regulation

$$\begin{aligned} \dot{S}_3(i_{1q}) = 0 \Rightarrow \\ u_{1qeq} = \sigma L_1 \dot{i}_{1qref} + R_{sm} i_{1q} + \sigma L_1 \omega_s i_{1d} + \frac{L_m}{L_2} \omega \lambda_{2d} \end{aligned} \quad (20)$$

$$\dot{S}_3(i_{1q}) S_3(i_{1q}) < 0 \Rightarrow u_{1qn} = K_q \cdot \text{sign}(S_3(i_{1q})) \quad (21)$$

The controller is given by

$$u_{1qc} = u_{1qeq} + u_{1qn} \quad (22)$$

To satisfy stability condition of the system, all of the following gains ($K_d, K_q, K_\phi, K_\omega$) should be chosen positive.

4. Load torque estimator

Since the load torque is not exactly known, its estimation is introduced. The mechanical equation gives

$$\hat{T}_L = \frac{pL_m}{L_2} \lambda_{2d} i_{1q} - J \frac{d\omega}{dt} - F\omega \quad (23)$$

5. Mras rotor resistance identification

As the estimated rotor flux is sensitive to rotor-resistance variation, a stable rotor-resistance MRAS estimator can be developed. We can rewrite the dynamic model of an induction motor given before by equations (11) as a compact form given by (Leonhard (1984); Pavlov & Zaremba (2001)) :

$$\frac{d\omega}{dt} = \frac{3}{2} \frac{n_p L_m}{J L_2} i_1^T M \lambda_2 - \frac{T_L}{J} \quad (24)$$

$$\frac{d\lambda_2}{dt} = \left(-\frac{R_2}{L_2} I + n_p \omega M \right) \lambda_2 + \frac{R_2}{L_2} J i_1 \quad (25)$$

$$\frac{di_1}{dt} = -\frac{J}{\sigma L_1 L_2} \left(-\frac{R_2}{L_2} I + n_p \omega M \right) \lambda_2 - \frac{1}{\sigma L_1} \left(R_1 + \frac{J^2 R_2}{L_2^2} \right) i_1 + \frac{1}{\sigma L_1} u_1 \quad (26)$$

Where

$$I = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}, \quad M = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix}$$

In order to design a rotor resistance identifier equations (25) (26) are transformed to eliminate the unobservable rotor flux. At this point an assumption is made that the rotor speed changes significantly slower relative to the rotor flux. Thus it may be treated as a constant parameter. First differentiating (26) and eliminating λ_2 gives

$$\frac{d^2 i_1}{dt^2} = (\alpha_1 I + \omega \beta_1 M) \frac{di_1}{dt} + (\alpha_2 I + \omega \beta_2 M) i_1 + (\alpha_3 I + \omega \beta_3 M) u_1 + \alpha_4 \frac{du_1}{dt} \quad (27)$$

where

$$\begin{aligned} \alpha_1 &= -\frac{1}{\sigma L_1} \left(R_1 + \frac{L_m^2 R_2}{L_2^2} \right) - \frac{R_2}{L_2}, & \beta_1 &= n_p, \\ \alpha_2 &= -\frac{R_2 R_1}{\sigma L_2 L_1}, & \beta_2 &= \frac{n_p R_1}{\sigma L_1}, \\ \alpha_3 &= \frac{R_2}{\sigma L_1 L_2}, & \beta_3 &= -\frac{n_p}{\sigma L_1}, & \alpha_4 &= \frac{1}{\sigma L_1}. \end{aligned}$$

Adding $c di_1/dt, c > 0$, to both sides of (27) and formally dividing by $(d/dt + c)$ transforms (27) to

$$\frac{di_1}{dt} = a + \omega b + \epsilon \quad (28)$$

where $\epsilon \rightarrow 0$ exponentially and the functions a and b are linear combinations of the filtered stator current and stator voltage command:

$$\begin{aligned} a &= (c + \alpha_1) i_{11} + \alpha_2 i_{10} + \alpha_3 u_{10} + \alpha_4 u_{11} \\ b &= M (\beta_1 i_{11} + \beta_2 i_{10} + \beta_3 u_{10}) \end{aligned}$$

where

$$\begin{aligned} i_{10} &= \frac{1}{s+c} i_1, & i_{11} &= \frac{s}{s+c} i_1 \\ u_{10} &= \frac{1}{s+c} u_1, & u_{11} &= \frac{s}{s+c} u_1 \end{aligned} \quad (29)$$

Here s denotes d/dt .

To obtain a reference model for the rotor resistance identification the part of the right-hand side of (28) containing R_2 is separated:

$$\frac{di_1}{dt} = f_1 + R_2 f_2 + \omega M f_3 + \epsilon \quad (30)$$

where

$$\begin{aligned} f_1 &= (c + \rho_1) i_{11} + \rho_2 u_{11} \\ f_2 &= \gamma_1 i_{11} + \gamma_2 i_{10} + \gamma_3 u_{10} \\ f_3 &= \beta_1 i_{11} + \beta_2 i_{10} + \beta_3 u_{10} \end{aligned} \quad (31)$$

Coefficients in (31) are calculated according to the following formulae:

$$\begin{aligned} \rho_1 &= -\frac{R_1}{\sigma L_1}, & \rho_2 &= \frac{1}{\sigma L_1}, \\ \beta_1 &= n_p, & \beta_2 &= \frac{n_p R_1}{\sigma L_1}, & \beta_3 &= -\frac{n_p}{\sigma L_1}, \\ \gamma_1 &= -\frac{1}{L_2} \left(1 + \frac{L_m^2}{\sigma L_1 L_2} \right), & \gamma_2 &= \frac{\rho_1}{L_2}, & \gamma_3 &= \frac{\rho_2}{L_2}. \end{aligned}$$

Since ω is available for measurement the design of an R_2 identifier is straightforward. It is based on the MRAS identification approach (Sastry & Bodson (1989)) with (30) being a reference model.

Within this approach an identifier consists of a tuning model depending on an estimate of the unknown parameter and a mechanism to adjust the estimate. This adjustment is performed to make the output of the tuning model asymptotically match the output of the reference model. In our case the tuning model is given by

$$\frac{d\hat{i}_1}{dt} = -L (\hat{i}_1 - i_1) + f_1 + \omega M f_3 + \hat{R}_2 f_2 \quad (32)$$

where $L > 0$ is a constant and \hat{R}_2 is the estimate of R_2 . The dynamics of the error $e = \hat{i}_1 - i_1$ is the following

$$\frac{de}{dt} = -Le + (\hat{R}_2 - R_2) f_2 - \epsilon \quad (33)$$

The adjustment equation for \hat{R}_2 is

$$\frac{d\hat{R}_2}{dt} = -\gamma (\hat{i}_1 - i_1)^T f_2, \quad \gamma > 0. \quad (34)$$

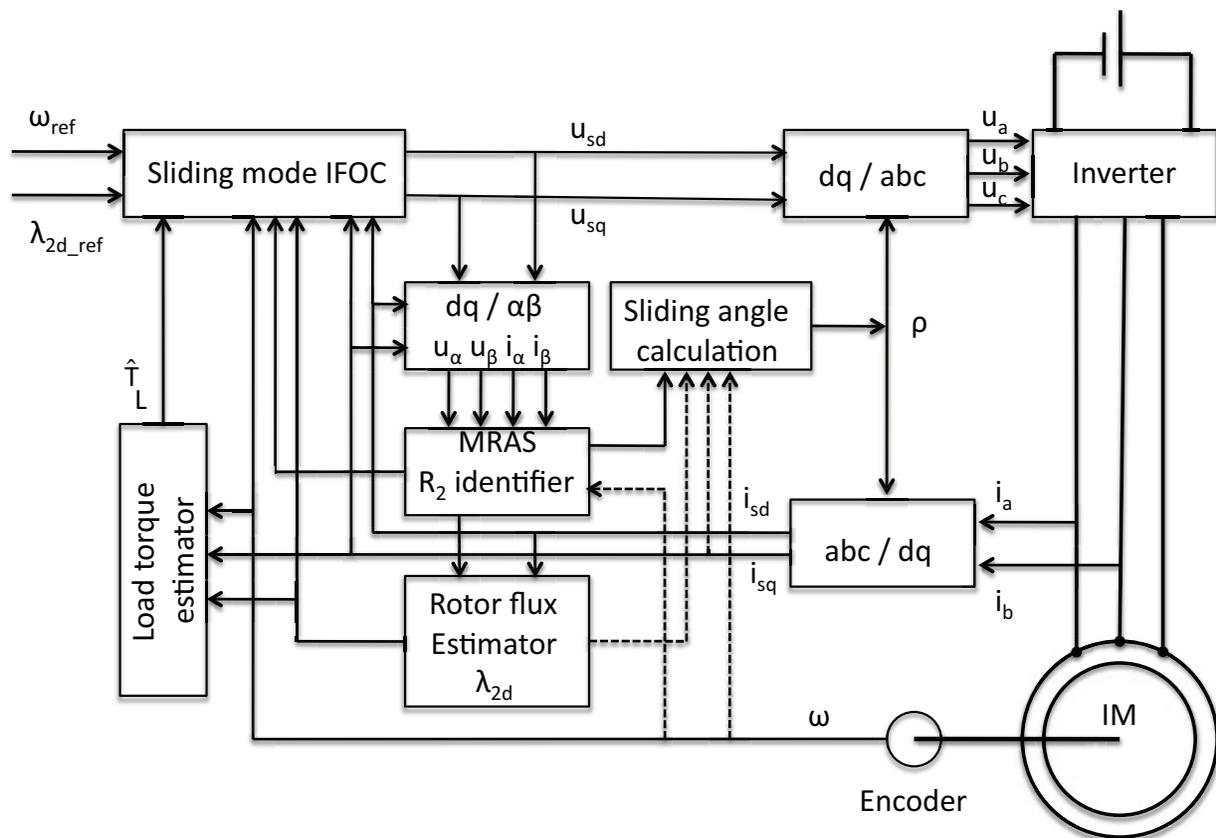


Fig. 2. Overall bloc diagram of the control scheme for IM

6. Simulations results

The overall configuration of the control system for IM is shown in Fig.2. The effectiveness of the proposed controller combined with the rotor resistance and load torque estimations has been verified by simulations in Matlab/Simulink. The parameters of the induction motor used are given in Appendix. The simulation results have been obtained under a constant load torque of 10 Nm.

Parameters of the Sliding mode controllers are: $K_d = 500$, $K_q = 500$, $K_\phi = 400$ and $K_\omega = 300$. Parameters of the MRAS identifier are: $\gamma = 0.2$, $L = 100$ and $c = 0.01$.

Results obtained are shown in Fig.3. The reference speed is set to 200 rad/s until $t=4s$, when it is reversed to -200rad/s to allow drive to operate in the generating mode. The reference flux is set to 0.4wb. The load torque is changed from 0 to 10 Nm at $t=0.6s$. The rotor resistance and the load torque estimators are activated. It can be seen that the estimated load torque converges very quickly to the actual value. In addition, the estimated values of R_2 follow its actual value very closely.

In order to show the convergence capability of the MRAS rotor resistance estimator, at $t=1.2s$, the interne value of R_2 in the MAS model has been disturbed and varied intentionally 30%

of its initial value and held constant until $t=2.5s$, at the same moment, the rotor resistance estimator is disconnected from the control structure.

It can be noted that the error in the estimated value of R_2 produces a steady state error in the speed and the rotor flux control and also generates an error in the estimated value of the load torque.

At $t=2.5s$, the initial value of R_2 in MAS model has restored and the rotor resistance estimator has been reconnected and the rotor resistance is seen to converge to the actual value and also the other system variables.

At $t=4s$, the reference speed is reversed to -200 rad/s to allow the drive to operate in the generating mode.

The results show a stable operation of the drive in the various operating modes.

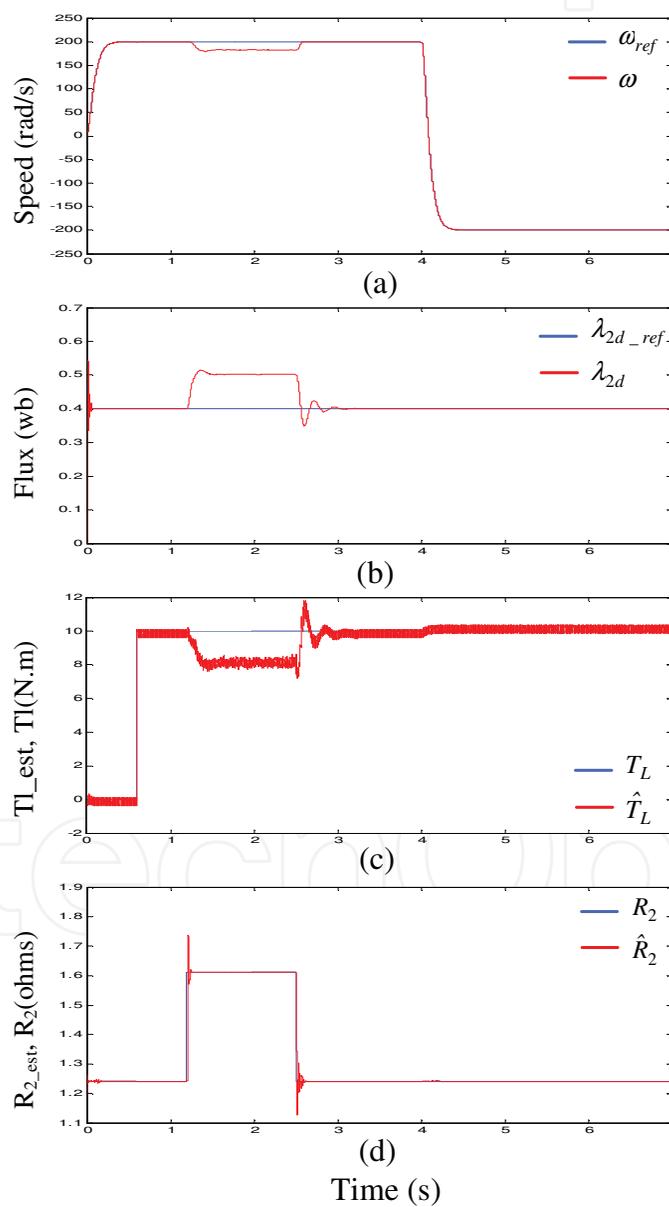


Fig. 3. Tracking Performance and parameters estimates : a)reference and actual motor speeds, b)reference and actual rotor flux, c) T_L and \hat{T}_L (N.m), d) R_2 and \hat{R}_2 (Ω).

Furthermore, simulation results have been performed to show the capability of the load torque estimator to track the rapid load torque changes and also to show the performance of flux and speed control. The results are shown in Fig. 4. These results have been obtained using the same parameters used for the results in Fig. 3. The rotor resistance and load torque estimators are activated at $t=0.2$ s. The load torque has been reversed from 10 Nm to -10 Nm at $t=2$ s. It can be seen from Fig. 4 that the estimated load torque converges rapidly to its actual value and the rotor resistance estimator is stable. In addition, the results in Fig. 4 show an excellent control of rotor flux and speed.

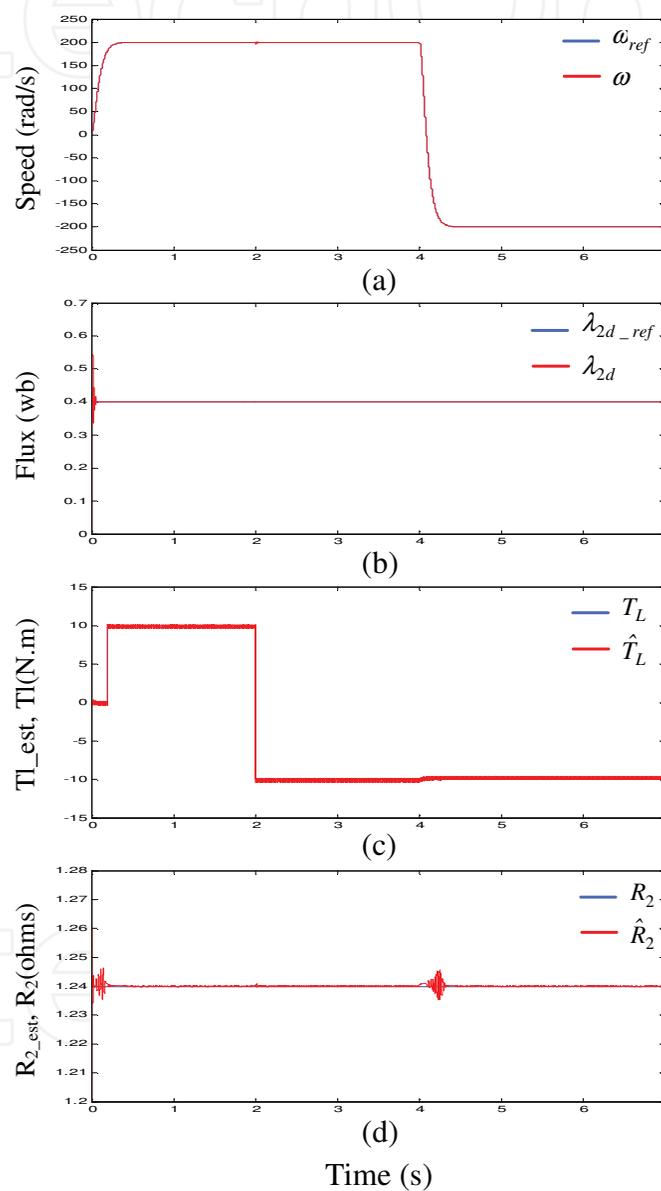


Fig. 4. Tracking Performance and parameters estimates : a)reference and actual motor speeds, b)reference and actual rotor flux, c) T_L and \hat{T}_L (N.m), d) R_2 and \hat{R}_2 (Ω).

Thus, the simulation results confirm the robustness of the proposed scheme with respect to the variation of the rotor resistance and load torque.

7. Conclusions and future works

In this paper, a novel scheme for speed and flux control of induction motor using online estimations of the rotor resistance and load torque have been described. The nonlinear controller presented provides voltage inputs on the basis of rotor speed and stator currents measurements and guarantees rapid tracking of smooth speed and rotor flux references for unknown parameters (rotor resistance and load torque) and non-measurable state variables (rotor flux). In simulation results, we have shown that the proposed nonlinear adaptive control algorithm achieved very good tracking performance within a wide range of the operation of the IM. The proposed method also presented a very interesting robustness properties with respect to the extreme variation of the rotor resistance and reversal of the load torque. The other interesting feature of the proposed method is that it is simple and easy to implement in real time.

From a practical point of view, in order to reduce the chattering phenomenon due to the discontinuous part of the controller, the $\text{sign}(\cdot)$ functions have been replaced by the saturation functions $\frac{(\cdot)}{(\cdot)+0.01}$ (Slotine & Li (1991)).

It would be meaningful in the future work to implement in real time the proposed algorithm in order to verify its robustness with respect to the discretization effects, parameter uncertainties and modelling inaccuracies.

| Induction motor data | | |
|----------------------|--------|----------|
| Stator resistance | 1.34 | Ω |
| Rotor resistance | 1.24 | Ω |
| Mutual inductance | 0.17 | H |
| Rotor inductance | 0.18 | H |
| Stator inductance | 0.18 | H |
| Number of pole pairs | 2 | H |
| Motor load inertia | 0.0153 | Kgm^2 |

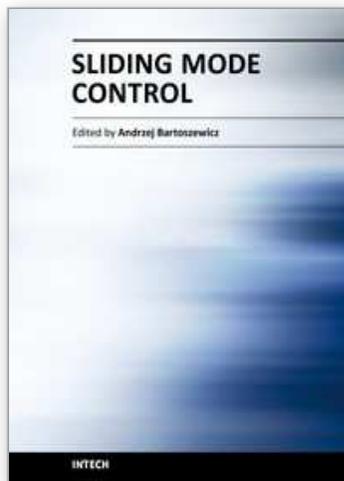
| Nomenclature | |
|------------------------|---------------------------------|
| R_1, R_2 | rotor, stator resistance |
| i_1, i_2 | rotor, stator current |
| λ_1, λ_2 | rotor, stator flux linkage |
| λ_{2d} | amplitude of rotor flux linkage |
| u_1, u_2 | rotor, stator voltage input |
| ω | rotor angular speed |
| ω_s | stator angular frequency |
| ρ | rotor flux angle |
| n_p | number of pole pairs |
| L_1, L_2 | rotor, stator inductance |
| T_r | rotor time constant |
| L_m | mutual inductance |
| J, T_L | inercia, load torque |
| F | coefficient of friction |
| $(\cdot)_d, (\cdot)_q$ | in (d,q) frame |
| $(\cdot)_a, (\cdot)_b$ | in (a,b) frame |
| $\hat{(\cdot)}$ | estimate of (\cdot) |
| $(\cdot)^*$ | reference of (\cdot) |

$$\sigma = 1 - \frac{L_m^2}{L_1 L_2}, \alpha = \frac{1}{L_2}, \beta = \frac{L_m}{\sigma L_1 L_2}, \eta_1 = \frac{R_1}{\sigma L_1}$$

$$\eta_2 = \frac{L_m^2}{\sigma L_1 L_2^2}, \mu = \frac{3n_p L_m}{2J L_2}, R_{sm} = R_1 + \frac{L_m^2}{L_2^2} R_2$$

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The main objective of this monograph is to present a broad range of well worked out, recent application studies as well as theoretical contributions in the field of sliding mode control system analysis and design. The contributions presented here include new theoretical developments as well as successful applications of variable structure controllers primarily in the field of power electronics, electric drives and motion steering systems. They enrich the current state of the art, and motivate and encourage new ideas and solutions in the sliding mode control area.

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