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SPEED SLIDING MODE CONTROL OF SENSORLESS INDUCTION MACHINE

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In this paper, a direct field-oriented induction motor drive with a sliding mode controller (SMC) is presented. The design includes rotor speed estimation from measured stator terminal voltages and currents. For that reason we have defined a simple speed estimator. The (SMC) technique finds its stronger justification in the utilization of a robust nonlinear control law to model uncertainties. The used control algorithm is a sliding mode associated with a particular function "sat" to limit chattering effects that presents a serious problem in applications to variable structure systems. Simulation tests under load disturbances and parameter uncertainties are provided to evaluate the consistency and performance of the proposed control technique.

K e y w o r d s: induction machine, motor drives, sliding mode control, sensorless machine

1 INTRODUCTION

The control of the induction machine (IM) must take into account machine specifics: the high order of the model, nonlinear functioning as well as the coupling between the different variables of control. Furthermore, the machine parameters depend generally on the operating point and vary either on the temperature (resistance), or with the magnetic state of the induction machine, without taking into account the variation. These parametric variations modify the performances of the control system when we use a regulator or a control law with fixed parameters. The new industrial applications necessitate speed variators having high dynamic performances, good precision in permanent regime, high capacity of overload in the whole range of speed and robustness to different perturbations. Thus, the recourse to robust control algorithms is desirable in stabilization and in tracking trajectories. The variable structure control (VSC) possesses this robustness using the sliding mode control that can offer many good properties such as good performance against unmodelled dynamics, insensitivity to parameter variation, external disturbance rejection and fast dynamic [5–11]. These advantages of sliding mode control can be employed in the position and speed control of an alternative current servo system.

Using field oriented control (FOC) of the induction machine, the knowledge of rotor speed and flux is necessary. Tachogenerators or digital shaft-position encoders are usually used to detect the rotor speed of motors. These speed sensors lower the system reliability and require special attention to noise. In addition, for some special applications such as very high-speed motor drives, there exist difficulties in mounting these speed sensors. Recently many researches have been carried on the design of speed sensorless control schemes. In this work the speed is obtained based on the measurement of stator voltages and currents. However, the estimation is usually complex and heavily dependent on machine parameters. Therefore, although sensorless vector-controlled drives are commercially available at this time, the parameter uncertainties impose a challenge in the control performances.

In this paper, we begin with the IM oriented model in view of the vector control, next the rotor flux Φ_r, the stator current I_{sq} and the mechanical speed Ω are estimated. We then present the sliding mode theory and design of the sliding mode rotor flux controller and motor speed. Finally, we give some conclusion remarks on the speed sensorless control proposed of IM using the sliding mode.

2 INDUCTION MOTOR ORIENTED MODEL

The model of a three phase squirrel cage induction motor in the synchronous reference frame whose axis d is aligned with the rotor flux vector, ($\Phi_{rd} = \Phi_r$ and $\Phi_{rq} = 0$), can be expressed as [1–4]

$$\dot{I}_{sd} = -\gamma_{sd} + \omega_s I_{sq} + \frac{K}{T_r}\Phi_{rd} + \frac{1}{\sigma L_s}U_{sd} \tag{1}$$

$$\dot{I}_{sq} = -\omega I_{sd} - \gamma I_{sq} - P\Omega K\Phi_{rd} + \frac{1}{\sigma L_s}U_{sq} \tag{2}$$

$$\dot{\Phi}_{rd} = M_{sr}I_{sd} - \frac{1}{T_r}\Phi_{rd} \tag{3}$$

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$$\dot{\Phi}_{rq} = \frac{M_{sr}}{T_r} I_{sq} - (\omega_s - P\Omega)\Phi_{rd} \tag{4}$$

$$\dot{\Omega} = \frac{PM_{sr}}{JL_r}(\Phi_{rd} I_{sq}) - \frac{C_r}{J} - f\Omega \tag{5}$$

with $T_r = \frac{L_r}{R}$, $\sigma = 1 - \frac{M_{sr}^2}{L_s L_r}$,

$$K = \frac{M_{sr}}{\sigma L_s L_r}, \quad \gamma = \frac{R_s}{\sigma L_s} + \frac{R_r M_{sr}^2}{\sigma L_s L_r^2}.$$

Here Φ_{rd}, Φ_{rq} are rotor flux components, U_{sd}, U_{sq} are stator voltage components, I_{sd}, I_{sq} are stator current components, σ is the leakage factor and p is the number of pole pairs. R_s and R_r are stator and rotor resistances, L_s and L_r denote stator and rotor inductances, whereas M_{sr} is the mutual inductance. T_e is the electromagnetic torque, C_r is the load torque, J is the moment of inertia of the IM, Ω is mechanical speed, ω_s is stator pulsation, f is damping coefficient, T_r is the rotoric time constant.

3 SPEED ESTIMATOR

In order to obtain an accurate dynamic representation of the motor speed, we take an algorithm of estimation based on the integration of the stator voltage equations in the stationary frame. The speed estimation Ω can be obtained by the next equation.

$$\widehat{\Omega} = \frac{1/P}{\widehat{\Phi}_{r\alpha}^2 + \widehat{\Phi}_{r\beta}^2} \Big(\widehat{\Phi}_{r\alpha} \dot{\widehat{\Phi}}_{r\beta} - \widehat{\Phi}_{r\beta} \dot{\widehat{\Phi}}_{r\alpha} - \frac{M_{sr}}{T_r} \big(\widehat{\Phi}_{r\alpha} I_{s\beta} - \widehat{\Phi}_{r\beta} I_{s\alpha} \Big) \tag{6}$$

with

$$\dot{\widehat{\Phi}}_{r\alpha} = \frac{L_r}{M_{sr}} \big(U_{s\alpha} - R_s I_{s\alpha} - \sigma L_s \dot{I}_{s\alpha} \,, \tag{7}$$

$$\dot{\widehat{\Phi}}_{r\beta} = \frac{L_r}{M_{sr}} \big(U_{s\beta} - R_s I_{s\beta} - \sigma L_s \dot{I}_{s\beta} \big) \,. \tag{8}$$

$\widehat{\Theta}_s$ is the angle between the rotoric vector flux Φ_r and the axis of the (α, β) frame

$$\widehat{\Theta}_s = \arctan\Big(\frac{\widehat{\Phi}_{r\beta}}{\widehat{\Phi}_{r\alpha}} \Big). \tag{9}$$

Here $\widehat{\Phi}_{r\alpha}, \widehat{\Phi}_{r\beta}$ are the estimated rotor flux components, $I_{r\alpha}, I_{r\beta}$ are the measured stator current components.

4 VARIABLE STRUCTURE CONTROL DESIGN

The sliding mode technique is developed from variable structure control (VSC) to solve the disadvantages of other designs of nonlinear control systems. The sliding mode is a technique to adjust feedback by previously defining a surface. The system which is controlled will be forced to that surface, then the behaviour of the system slides to the desired equilibrium point.

The main feature of this control is that we only need to drive the error to a "switching surface". When the system is in "sliding mode", the system behaviour is not affected by any modelling uncertainties and/or disturbances.

The design of the control system will be demonstrated for a nonlinear system presented in the canonical form [5–9]

$$\dot{x} = f(x,t) + B(x,t)u(x,t) \,, \quad x \in \Re^n, \; u \in \Re^m, \quad \operatorname{rank}\big(B(x,t)\big) = m \tag{10}$$

with control in the sliding mode, the goal is to keep the system motion on the manifold S, which is defined as

$$S = \{x \colon e(x,t) = 0\} \,, \tag{11}$$

$$e = x^d - x \,. \tag{12}$$

Here e is the tracking error vector, x^d is the desired state vector, x is the state vector. The control input u has to guarantee that the motion of the system described in (10) is restricted to belong to the manifold S in the state space. The sliding mode control should be chosen such that the candidate Lyapunov function satisfies the Lyapunov stability criteria:

$$V = \frac{1}{2} S(x)^2, \tag{13}$$

$$\dot{V} = S(x)\, \dot{S}(x) \,. \tag{14}$$

This can be assured for

$$\dot{V} = -\eta \big| S(x) \big| \,. \tag{15}$$

Here η is strictly positive. Essentially, equation (13) states that the squared "distance" to the surface, measured by $e(x)^2$, decreases along all system trajectories. Therefore (14), (15) satisfy the Lyapunov condition. With the selected Lyapunov function the stability of the whole control system is guaranteed. The control function will satisfy reaching conditions in the following form:

$$u^{com} = u^{eq} + u^n \,. \tag{16}$$

Here u^{com} is the control vector, u^{eq} is the equivalent control vector, u^n is the correction factor and must be calculated so that the stability conditions for the selected control are satisfied.

$$u^n = K \operatorname{sat}\big(S(x)/\phi \big) \tag{17}$$

$\operatorname{sat}\big(S(x)/\phi\big)$ is the proposed saturation function, ϕ is the boundary layer thickness. In this paper we propose the Slotine method

$$S(X) = \Big(\frac{\mathrm{d}}{\mathrm{d}t} + \lambda \Big)^{n-1} e \,. \tag{18}$$

Here, e is the tracking error vector, λ is a positive coefficient and n is the system order.

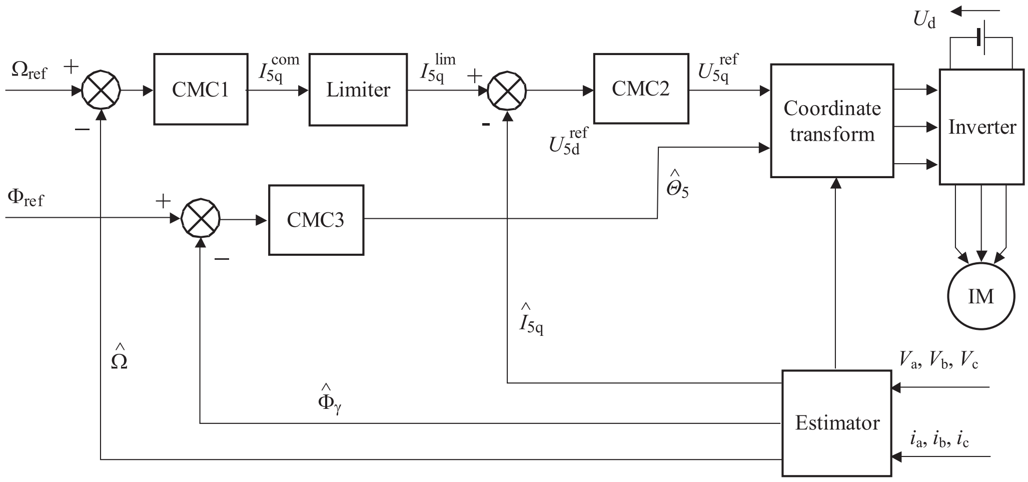

Fig. 1. Block diagram of the proposed control scheme of IM

4.1 Speed control

To control the speed of the induction machine, the sliding surface defined by equation (13) for $n = 1$ can be obtained as:

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

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

Substituting the expression of $\dot{\Omega}$ equation (1) in equation (18), we obtain

$$\dot{S}(\Omega) = \dot{\Omega}_{ref} - \frac{PM_{sr}}{JL_r}\big(\Phi_{rd}I_{sq}\big) - \frac{C_r}{J} - f\Omega. \tag{21}$$

We take

$$I_{sq} = I_{sq}^{eq} + I_{sq}^{n}. \tag{22}$$

- During the sliding mode and in permanent regime, we have $S(\Omega) = 0$, $\dot{S}(\Omega) = 0$, where the equivalent control

$$I_{sq}^{eq} = \frac{JL_r}{PM_{sr}\Phi_{rd}}\Big(\dot{\Omega}_{ref} + \frac{f}{J}\Omega + \frac{C_r}{J}\Big)\,. \tag{23}$$

- During the convergence mode, the condition $S(\Omega)\dot{S}(\Omega) \le 0$ must be verified. We obtain

$$\dot{S}(\Omega) = -\frac{P^2 M_{sr}\Phi_{rd}}{JL_r} I_{sq}^{n}\,. \tag{24}$$

Therefore, the correction factor is given by

$$I_{sq}^{n} = K_{isq}\,\mathrm{sat}\big(S(\Omega)\big)\,. \tag{25}$$

To verify the system stability condition, parameter K_{isq} must be positive.

4.2. Isq stator current control

In order to limit all possible overshoot of the current I_{sq}, we add a limiter of current defined by

$$I_{sq}^{\lim} = I_{sq}^{\max}\,\mathrm{sat}\big(I_{sq}^{com}\big)\,. \tag{26}$$

The current control manifold is

$$S\big(I_{sq}\big) = I_{sq}^{\lim} - I_{sq}\,. \tag{27}$$

The control voltage is

$$U_{sq}^{ref} = U_{sq}^{eq} + U_{sq}^{n} \tag{28}$$

$$U_{sq}^{eq} = \sigma L_s\big(\dot{I}_{sq}^{\lim} + \omega_s I_{sd} + \gamma I_{sq} + P\Omega K\Phi_{rd}\big) \tag{29}$$

$$U_{sq}^{n} = Ku_{sq}\,\mathrm{sat}\big(S(I_{sq})\big)\,. \tag{30}$$

To verify the system stability condition, parameter K_{uqs} must be positive.

4.3 Flux control

In order to appear control U_{sd}, we take $n = 2$, the manifold equation can be obtained by

$$S(\Phi_r) = \lambda_\Phi\big(\Phi_r^{ref} - \Phi_r\big) + \big(\dot{\Phi}_r^{ref} - \dot{\Phi}_r\big) \tag{31}$$

the control voltage is

$$U_{sd} = U_{sd}^{eq} + U_{sd}^{n} \tag{32}$$

$$U_{sd}^{eq} = -\sigma L_s\Bigg(\Big(\dot{\Phi}_r^{ref} + \lambda_\Phi\dot{\Phi}_r^{ref} + \Big(\frac{1}{T_r} - \lambda_\Phi\Big)\dot{\Phi}_r\Big)\frac{T_r}{M_{sr}} - \Big(-\gamma_{sd} + \omega_s I_{sq} + \frac{K}{T_r}\Phi_{rd}\Big)\Bigg) \tag{33}$$

$$U_{sd}^{n} = Ku_{sd}\,\mathrm{sat}\big(S(\Omega)\big)\,. \tag{34}$$

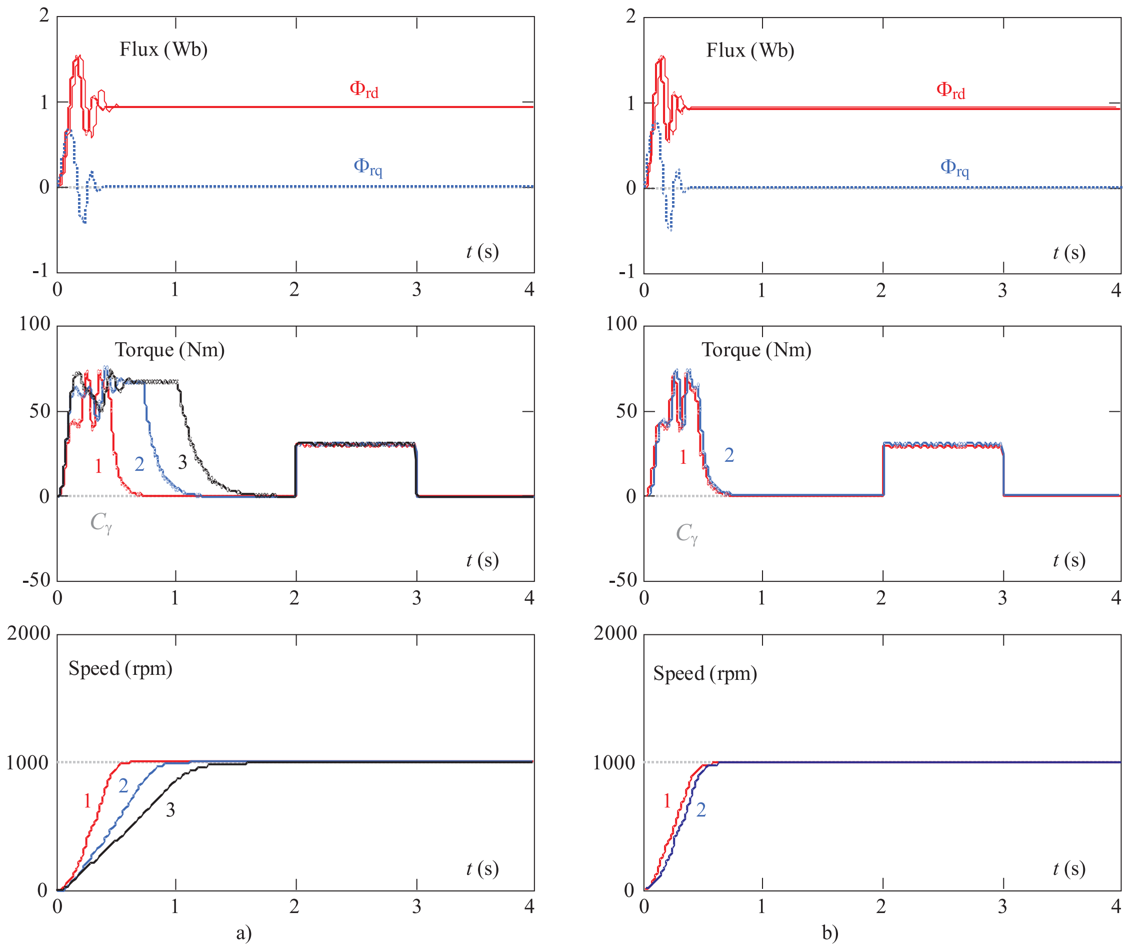

Fig. 2. System responses (a) $J = mJ_n$, $1 \leq m \leq 3$ (b) 1: Nominal case, 2: an increase of $0.5\,R_s$, $0.5\,R_r$, $0.2\,L_s$, $0.2\,L_r$ and $0.2\,M_{sr}$.

To verify the system stability condition, parameter Ku_{sd} must be positive. The selection of coefficients Ki_{sq}, Ku_{sd}, Ku_{sq} and λ_Φ must be made to satisfy following requirements:

- Existence condition of the sliding mode, which requires that the state trajectories are directed toward the sliding manifold,
- Hitting condition, which requires that the system trajectories encounter the manifold sliding irrespective of their starting point in the state space(insure the rapidity of the convergence),
- Stability of the system trajectories on the sliding manifold,
- Not saturate the control to allow the application of the control discontinuous.

5 SIMULATION RESULTS

5.1 System description

The block diagram of the proposed robust control scheme is presented in Fig. 1. The blocks SMC1, SMC2, SMC3 are sliding mode controllers which represent, respectively, the speed controller, the current controller and the flux controller. The block limiter limits the current within the limit value. The block 'coordinate transform' makes the conversion between the synchronously rotating and stationary reference frame. The IM is fed by a voltage inverter. The block Ω, I_{sq} and Φ_r are estimated by the bloc 'Estimator'. The IM used in this work is a 7.5 kW, $U = 220$ V, 50 Hz, $P = 2$, $I_n = 16$ A, $\Phi_n = 0.9$ Wb. IM parameters: $R_s = 0.63\,\Omega$, $R_r = 0.4\,\Omega$, $L_r = M_{sr} = 0.097$ H, $L_s = 0.091$ H. The system has the following mechanical parameters: $J = 0.22\,\text{kgm}^2$, $f = 0.001$ Nms/rd.

The global system is simulated in real time by the software Matlab/Simulink.

5.2 Simulation results

To illustrate performances of control, we simulated a loadless starting up mode with the reference speed +1000 rpm and an application and elimination of the load torque ($C_r = 30$ Nm) at time 2 s and 3 s.

In order to test the robustness of the proposed control, we have studied the speed performances with current limitation. The introduced variations in tests look in practice to work conditions as the magnetic circuit overheating and saturation. Two cases are considered:

1. Inertia variation,
2. Stator and rotor resistance variations with stator and rotor inductance and mutual variations.

The Fig.2 shows the tests of the robustness: a) The robustness tests in relation to inertia variations. b) The robustness tests concerning the variation of the resistances and the inductances in the stator and the rotor.

From the simulation results, it appears that the parameter variation does not allocate the performances of the proposed control. The flux tracks the desired flux and it is insensitive to external and internal parameter variations of the machine. The decoupling between the flux and torque is ensured. The started torque is limited and the speed response stays insensitive to parameter variations of the machine, without overshoot and without static error, the perturbation reject is instantaneous.

Figure 2b shows the parameter variation does not allocate performances of proposed control. The speed response is insensitive to parameter variations of the machine, without overshoot and without static error. The other performances are maintained.

6 CONCLUSION

In this paper a speed sliding mode control of a sensorless induction machine using the field oriented control has been presented. An algorithm to estimate the rotor speed, current I_{sq} (torque) and rotor flux is presented. Results of simulation show the robustness of the proposed control in relation to the presence of internal and external perturbations. Decoupling of the torque and flux of IM is guaranteed. The rotor flux tracks the reference value. With a good choice of the parameters of control and smoothing out control discontinuity, the chattering effects are reduced and torque fluctuations are decreased. The speed tracking is without overshoot and zero static error. Furthermore, this regulation presents a simple robust control algorithm that has the advantage to be easily implantable in a calculator.

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