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INTELLIGENT ROBUST CONTROL LAW FOR INDUCTION MOTORS BASED ON FIELD ORIENTED CONTROL THEORY

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ABSTRACT

In this paper a sensorless adaptive robust control law is proposed to improve the trajectory tracking performance of induction motors. The proposed design employs the so called vector (or field oriented) control theory for the induction motor drives and the designed control law is based on an integral sliding-mode algorithm that overcomes the system uncertainties. The proposed sliding-mode control law incorporates an adaptive switching gain to avoid calculating an upper limit of the system uncertainties. The proposed design also includes a new method in order to estimate the rotor speed. In this method, the rotor speed estimation error is presented as a first order simple function based on the difference between the real stator currents and the estimated stator currents. The stability analysis of the proposed controller under parameter uncertainties and load disturbances is provided using the Lyapunov stability theory.

KEY WORDS

Robust Control, Adaptive Control, Power Systems, Nonlinear Systems, Modelling and Simulation.

1 Introduction

Field oriented control method is widely used for advanced control of induction motor drives. By providing decoupling of torque and flux control demands, the vector control can govern an induction motor drive similar to a separate excited direct current motor without sacrificing the quality of the dynamic performance. However, the field oriented control of induction motor drives presents two main problems that have been providing quite a bit research interest in the last decade. The first one relies on the uncertainties in the machine models and load torque, and the second one is the precise computation of the motor speed without using speed sensors.

The decoupling characteristics of the vector control is sensitive to machine parameters variations. Moreover, the machine parameters and load characteristics are not exactly known, and may vary during motor operations. To overcome the above system uncertainties, the variable structure control strategy using the sliding-mode has been focussed on many studies and research for the control of the AC servo drive system in the past decade [2], [6]. However the traditional sliding control schemes requires the prior knowledge of an upper bound for the system uncertainties since this bound is employed in the switching gain calculation. This upper bound should be determined as precisely as possible, because as higher is the upper bound higher value should be considered for the sliding gain, and therefore the control effort will also be high, which is undesirable in a practice. In order to surmount this drawback, in the present paper it is proposed an adaptive law to calculate the sliding gain which avoids the necessity of calculate an upper bound of the system uncertainties.

Otherwise, a suitable speed control of an induction motor requires a precise speed information, therefore, a speed sensor, such a resolver and encoder, is usually adhered to the shaft of the motor to measure the motor speed. However, a speed sensor can not be mounted in some cases, such as motor drives in adverse environments, or high-speed motor drives. Therefore, sensorless induction motor drives are widely used in industry for their reliability and flexibility, particularly in hostile environments. Speed estimation methods using Model Reference Adaptive System MRAS are the most commonly used as they are easy to design and implement [3]. However, the performance of these methods is deteriorated at low speed because of the increment of nonlinear characteristics [5], [6].

In this paper the authors proposes a robust sensorless vector control scheme consisting on the one hand of an adaptive rotor speed estimation method based on MRAS in order to improve the performance of a sensorless vector controller in a low speed region. The proposed method can provide a fast speed estimation and improve the performance of other speed estimation methods in a low speed region and at zero-speed. This paper is organized as follows. The rotor speed estimation is introduced in Section 2. In section 3, the proposed robust speed control with adaptative sliding gain is presented, and in section 4 it is proposed a continuous approximation of the control law. Then the closed loop stability of the proposed scheme is demonstrated using the Lyapunov stability theory, and the exponential convergence of the controlled speed is also provided. In the Section 5, some simulation results are presented. Finally some concluding remarks are stated in the last Section.

2 Proposed rotor speed estimator

Many schemes [7], based on simplified motor models have been devised to sense the speed of the induction motor from measured terminal quantities for control purposes. In order to obtain an accurate dynamic representation of the motor speed, it is necessary to base the calculation on the coupled circuit equations of the motor. However, the performance of these methods is deteriorated at a low speed because of the increment of nonlinear characteristic of the system [6].

The current paper proposes a new rotor speed estimation method to improve the performance of a sensorless vector controller in the low speed region and at zero speed.

Since the motor voltages and currents are measured in a stationary frame of reference, it is also convenient to express these equations in that stationary frame.

From the stator voltage equations in the stationary frame it is obtained [4]:

$$\dot{\psi}_{dr} = \frac{L_r}{L_m}\left[v_{ds} - R_s i_{ds} - \sigma L_s \frac{d}{d\,t} i_{ds}\right] \qquad (1)$$

$$\dot{\psi}_{qr} = \frac{L_r}{L_m}\left[v_{qs} - R_s i_{qs} - \sigma L_s \frac{d}{d\,t} i_{qs}\right] \qquad (2)$$

where ψ is the flux linkage; L is the inductance; v is the voltage; R is the resistance; i is the current and $\sigma = 1 - L_m^2/(L_r L_s)$ is the motor leakage coefficient. The subscripts r and s denotes the rotor and stator values respectively refereed to the stator, and the subscripts d and q denote the dq-axis components in the stationary reference frame.

Using the rotor flux and motor speed, the stator current is represented as:

$$i_{ds} = \frac{1}{L_m}\left[\psi_{dr} + w_r\, T_r\, \psi_{qr} + T_r\, \dot{\psi}_{dr}\right] \qquad (3)$$

$$i_{qs} = \frac{1}{L_m}\left[\psi_{qr} - w_r\, T_r\, \psi_{dr} + T_r\, \dot{\psi}_{qr}\right] \qquad (4)$$

where w_r is the rotor electrical speed and $T_r = L_r/R_r$ is the rotor time constant.

From the equations (3) and (4) and using the estimated speed, the stator current is estimated as:

$$\hat{i}_{ds} = \frac{1}{L_m}\left[\psi_{dr} + \hat{w}_r\, T_r\, \psi_{qr} + T_r\, \dot{\psi}_{dr}\right] \qquad (5)$$

$$\hat{i}_{qs} = \frac{1}{L_m}\left[\psi_{qr} - \hat{w}_r\, T_r\, \psi_{dr} + T_r\, \dot{\psi}_{qr}\right] \qquad (6)$$

where $\hat{i}ds$ and $\hat{i}qs$ are the estimated stator currents and $\hat{w}_r$ is the estimated rotor electrical speed.

Subtracting the equations of the estimated stator currents (5) and (6) from the equations of the stator currents (3) and (4) the difference in the stator current is obtained as:

$$i_{ds} - \hat{i}_{ds} = \frac{T_r}{L_m}\psi_{qr}\,(w_r - \hat{w}_r) \qquad (7)$$

$$i_{qs} - \hat{i}_{qs} = -\frac{T_r}{L_m}\psi_{dr}\,(w_r - \hat{w}_r) \qquad (8)$$

In the above equations (7) and (8), the difference of the stator current and the estimated stator current is a sinusoidal value because it is a function of the rotor flux. However, if equation (7) is multiplied by ψ_{qr} and equation (8) is multiplied by ψ_{dr} and then are added together it is obtained:

$$(i_{ds} - \hat{i}_{ds})\psi_{qr} - (i_{qs} - \hat{i}_{qs})\psi_{dr} = \frac{T_r}{L_m}\psi_{qr}\,(w_r - \hat{w}_r)\,(\psi_{dr}^2 + \psi_{dr}^2) \qquad (9)$$

Unlike the equations (7) and (8), equation (9) uses the rotor flux magnitude which remains constant. From equation (9) the error of the rotor speed is obtained as follows:

$$e_{w_r} = w_r - \hat{w}_r = c\left[(i_{ds} - \hat{i}_{ds})\psi_{qr} - (i_{qs} - \hat{i}_{qs})\psi_{dr}\right] \qquad (10)$$

where:

$$c = \frac{L_m}{T_r}\frac{1}{\psi_{dr}^2 + \psi_{qr}^2} = \frac{L_m}{T_r\,\psi_r^2}$$

Therefore, from the equation (10) the speed estimation error is calculated from the stator current and rotor flux.

Using Lyapunov stability theory we can derive the following adaptation law for speed estimation:

$$\frac{d\hat{w}_r}{dt} = \alpha\, e_{w_r}\,, \quad \alpha > 0 \qquad (11)$$

where α is de adaptation gain that should be chosen greater than zero.

To demonstrate that the previous adaptation law makes the estimated speed error drops into zero, we can define the following Lyapunov function candidate:

$$V(t) = \frac{1}{2}\, e_{w_r}^2(t)$$

On the basis of the fact that the velocity of outer control loop is much slower than the estimated inner loop, hence the assumption of w_r approaching a constant is reasonable on deriving the following equations. Then, the time derivative of the previous Lyapunov function candidate is:

$$\begin{aligned} \dot{V}(t) &= e_{w_r}\dot{e}_{w_r} \\ &= e_{w_r}(-\dot{\hat{w}}_r) \\ &= -\alpha\, e_{w_r}^2 \end{aligned} \qquad (12)$$

Using the Lyapunov's direct method, since $V(t)$ is clearly positive-definite, $\dot{V}(t)$ is negative definite and $V(t)$ tends to infinity as $e_{w_r}(t)$ tends to infinity, then the equilibrium at the origin $e_{w_r}(t) = 0$ is globally asymptotically stable. Therefore $e_{w_r}(t)$ tends to zero as the time t tends to infinity.

Moreover, taking into account the previous Lyapunov function we can conclude that the rotor speed error converges to zero exponentially. From equation (12) we can obtain that V derivative verifies:

$$\dot{V}(t) = -\alpha\, e_{w_r}^2 = -\frac{\alpha}{2} V(t) \qquad (13)$$

The solution of the previous differential equation is:

$$V(t) = \frac{1}{2}\, e_{w_r}^2(t) = V(t_0)\exp(-\frac{\alpha}{2}t)$$

which implies that the rotor speed error converges to zero exponentially.

Therefore, the rotor speed w_r can be calculated using the proposed speed estimator which only make use of the measured stator voltages and currents in order to estimate the rotor speed.

3 Variable structure robust speed control with adaptive sliding gain

In general, the mechanical equation of an induction motor can be written as:

$$J\dot{w}_m + Bw_m + T_L = T_e \qquad (14)$$

where J and B are the inertia constant and the viscous friction coefficient of the induction motor system respectively; T_L is the external load; w_m is the rotor mechanical speed in angular frequency, which is related to the rotor electrical speed by $w_m = 2\, w_r/p$ where p is the pole numbers and T_e denotes the generated torque of an induction motor, defined as [4]:

$$T_e = \frac{3p}{4}\frac{L_m}{L_r}(\psi_{dr}^e i_{qs}^e - \psi_{qr}^e i_{ds}^e) \qquad (15)$$

where ψ_{dr}^e and ψ_{qr}^e are the rotor-flux linkages, with the subscript 'e' denoting that the quantity is refereed to the synchronously rotating reference frame; i_{qs}^e and i_{ds}^e are the stator currents, and p is the pole numbers.

The angular position of the rotor flux vector ($\bar{\psi}_r$) related to the d-axis of the stationary reference frame may be calculated by means of the rotor flux components in this reference frame (ψ_{dr}, ψ_{qr}) as follows:

$$\theta_e = \arctan\left(\frac{\psi_{qr}}{\psi_{dr}}\right) \qquad (16)$$

where θ_e is the angular position of the rotor flux vector.

Using the field-orientation control principle [4] the current component i_{ds}^e is aligned in the direction of the rotor flux vector $\bar{\psi}_r$, and the current component i_{qs}^e is aligned in the direction perpendicular to it. At this condition, it is satisfied that:

$$\psi_{qr}^e = 0, \qquad \psi_{dr}^e = |\bar{\psi}_r| \qquad (17)$$

Therefore, taking into account the previous results, the equation of induction motor torque (15) is simplified to:

$$T_e = \frac{3p}{4}\frac{L_m}{L_r}\psi_{dr}^e i_{qs}^e = K_T\, i_{qs}^e \qquad (18)$$

where K_T is the torque constant, and is defined as follows:

$$K_T = \frac{3p}{4}\frac{L_m}{L_r}\psi_{dr}^{e^*} \qquad (19)$$

where $\psi_{dr}^{e^*}$ denotes the command rotor flux.

With the above mentioned proper field orientation, the dynamic of the rotor flux is given by [4]:

$$\frac{d\psi_{dr}^e}{d\,t} + \frac{\psi_{dr}^e}{T_r} = \frac{L_m}{T_r}\, i_{ds}^e \qquad (20)$$

Then, the mechanical equation (14) becomes:

$$\dot{w}_m + a\, w_m + f = b\, i_{qs}^e \qquad (21)$$

where the parameters are defined as:

$$a = \frac{B}{J}, \quad b = \frac{K_T}{J}, \quad f = \frac{T_L}{J}; \qquad (22)$$

Now, we are going to consider the previous mechanical equation (21) with uncertainties as follows:

$$\dot{w}_m = -(a + \triangle a)w_m - (f + \triangle f) + (b + \triangle b)i_{sq} \quad (23)$$

where the terms $\triangle a$, $\triangle b$ and $\triangle f$ represents the uncertainties of the terms a, b and f respectively. It should be noted that these uncertainties are unknown, and that the precise calculation of its upper bound are, in general, rather difficult to achieve.

Let us define the tracking speed error as follows:

$$e(t) = \hat{w}_m(t) - w_m^*(t) = w_m(t) - w_m^*(t) - \tilde{w}_m(t) \quad (24)$$

where w_m^* is the rotor speed command and $\tilde{w}_m = w_m - \hat{w}_m$ is the rotor speed estimation error, that (as it is demonstrated in section II) converges to zero exponentially.

Taking the derivative of the previous equation with respect to time yields:

$$\dot{e}(t) = \dot{w}_m(t) - \dot{w}_m^*(t) - \dot{\tilde{w}}_m(t) = -a\, e(t) + u(t) + d(t) \qquad (25)$$

where the terms collected in the signals $u(t)$ and $d(t)$ are:

$$u(t) = b\, i_{sq}(t) - a\, w_m^*(t) - f(t) - \dot{w}_m^*(t) \qquad (26)$$

$$d(t) = -\triangle a\, w_m(t) - \triangle f(t) + \triangle b\; i_{qs}^e(t) - \dot{\tilde{w}}_m(t) \quad (27)$$

To compensate for the above described uncertainties that are presented in the system, it is proposed a sliding adaptive control scheme. In the sliding control theory, the switching gain must be constructed so as to attain the sliding condition [9]. In order to meet this condition a suitable choice of the sliding gain should be made to compensate for the uncertainties. For selecting the sliding gain vector, an upper bound of the parameter variations, unmodelled dynamics, noise magnitudes, etc. should be known, but in practical applications there are situations in which these bounds are unknown, or at least difficult to calculate. A solution could be to choose a sufficiently high value for the sliding gain, but this approach could cause a to high control signal, or at least more activity control than it is necessary in order to achieve the control objective.

One possible way to overcome this difficulty is to estimate the gain and to update it by some adaptation law, so that the sliding condition is achieved.

Now, we are going to propose the sliding variable $S(t)$ with an integral component as:

$$S(t) = e(t) + \int_0^t (a+k)e(\tau)\,d\tau \tag{28}$$

where k is a constant gain, and a is a parameter that was already defined in equation (22).

Then the sliding surface is defined as:

$$S(t) = e(t) + \int_0^t (a+k)e(\tau)\,d\tau = 0 \tag{29}$$

Now, we are going to design a variable structure speed controller, that incorporates an adaptive sliding gain, in order to control the AC motor drive.

$$u(t) = -k\,e(t) - \hat{\beta}(t)\gamma\,\mathrm{sgn}(S) \tag{30}$$

where the k is the gain defined previously, $\hat{\beta}$ is the estimated switching gain, γ is a positive constant, S is the sliding variable defined in eqn. (28) and $\mathrm{sgn}(\cdot)$ is the signum function.

The switching gain $\hat{\beta}$ is adapted according to the following updating law:

$$\dot{\hat{\beta}} = \gamma\,|S| \qquad\qquad \hat{\beta}(0) = 0 \tag{31}$$

where γ is a positive constant that let us choose the adaptation speed for the sliding gain.

In order to obtain the speed trajectory tracking, the following assumptions should be formulated:

($\mathcal{A}\,1$) The gain k must be chosen so that the term $(a+k)$ is strictly positive. Therefore the constant k should be $k > -a$.

($\mathcal{A}\,2$) There exits an unknown finite non-negative switching gain β such that

$$\beta > d_{max} + \eta \qquad \eta > 0$$

where $d_{max} \geq |d(t)| \quad \forall\, t$ and η is a positive constant.

Note that this condition only implies that the uncertainties of the system are bounded magnitudes.

($\mathcal{A}\,3$) The constant γ must be chosen so that $\gamma > 0$.

Theorem 1 *Consider the induction motor given by equation (23). Then, if assumptions ($\mathcal{A}\,1$), ($\mathcal{A}\,2$) and ($\mathcal{A}\,3$) are verified, the control law (30) leads the rotor mechanical speed so that the speed error* $\mathrm{e}(t) = \hat{w}_m(t) - w_m^*(t) = w_m(t) - w_m^*(t) - \tilde{w}_m(t)$ *tends to zero as the time tends to infinity, and then* $w_m(t)$ *tends to* $w_m^*(t)$.

The proof of this theorem will be carried out using the Lyapunov stability theory [2].

Finally, the torque current command, $i_{sq}^*(t)$, can be obtained directly substituting eqn. (30) in eqn. (26):

$$i_{sq}^*(t) = \frac{1}{b}\left[k\,e - \hat{\beta}\gamma\,\mathrm{sgn}(S) + a\,w_m^* + \dot{w}_m^* + f\right] \tag{32}$$

Therefore, the proposed variable structure speed control with adaptive sliding gain resolves the speed tracking problem for the induction motor, with some uncertainties in mechanical parameters and load torque.

4 Continuous approximation of switching control law

A frequently encountered problem in sliding control is that the control signal given by eqn.(30) is not smooth since the sliding control law is discontinuous across the sliding surfaces, which causes the chattering phenomenon. Chattering is undesirable in practice, since it involves high control activity and further may excite high-frequency dynamics. This situation can be avoided by smoothing out the control chattering within a thin boundary layer of thickness $\xi > 0$ neighboring the switching surface [1], [8] . On the other hand, it is well known that when in an adaptive control system the signals are not persistently exciting the parameter drift phenomenon may appear [8]. In these situations many different strategies can be applied, from complex methods to confer self-excitation capability to the system without the presence of external exciting signals, to simpler approaches based on the use of dead zones, as it will be done in this paper to avoid the possible parameter drift phenomenon that may appear in the proposed sliding gain adaptation law (eqn. 31)

In this way, some modifications should be done in the control law (30) and in the sliding gain adaptation law (31), to overcome the above mentioned problems:

i) In order to smooth the control law (30), the sign function included in it is replaced by a saturation function, so that it becomes:

$$u(t) = -k\,e(t) - \hat{\beta}(t)\gamma\,\mathrm{sat}\left(\frac{S}{\xi}\right) \tag{33}$$

where the saturation function $sat(\cdot)$ is defined in the usual way:

$$\mathrm{sat}\left(\frac{S}{\xi}\right) = \begin{cases} \mathrm{sgn}(S) & if \quad |S| > \xi \\ \dfrac{S}{\xi} & otherwise. \end{cases}$$

and ξ represents the thickness of the boundary layer neighboring the switching surface.

ii) In order to avoid the parameter drift phenomenon, the sliding gain adaptation law is modified to:

$$\dot{\hat{\beta}} = \gamma\,|S_o| \qquad \hat{\beta}(0) = 0 \tag{34}$$

where S_o is defined by:

$$S_o = S - \xi\,\mathrm{sat}\left(\frac{S}{\xi}\right)$$

It is interesting to point out that S_o is a measure of the distance from the sliding surface S to the interval $[-\xi,\ \xi]$:

$$S_o = \begin{cases} S - \xi & if \quad |S| > \xi \\ 0 & otherwise. \end{cases} \tag{35}$$

From the previous equation it is concluded that $\dot{S}_o = \dot{S}$ when S is outside the interval $[-\xi,\ \xi]$, while $\dot{S}_o = 0$ otherwise.

Theorem 2 *Consider the induction motor given by equation (23). Then, if assumptions $(\mathcal{A}1)$, $(\mathcal{A}2)$ and $(\mathcal{A}3)$ are verified, the control law (33) leads the rotor mechanical speed so that the speed tracking error $e(t) = \hat{w}_m(t) - w_m^*(t)$ can be made as small as desired by choosing an adequately small boundary layer thickness ξ.*

The proof of this theorem will be carried out using the Lyapunov stability theory.

Proof : Let us define the following Lyapunov function candidate:

$$V(t) = \frac{1}{2}S_o(t)S_o(t) + \frac{1}{2}\tilde{\beta}(t)\tilde{\beta}(t) \tag{36}$$

whose time derivative is given by:

$$\begin{aligned}
\dot{V}(t) = \;& S_o(t)\dot{S}(t) + \tilde{\beta}(t)\dot{\tilde{\beta}}(t) \\
&= S_o \cdot [\dot{e} + (a+k)e] + \tilde{\beta}(t)\dot{\hat{\beta}}(t) \\
&= S_o \cdot [(-a\,e + u + d) + (k\,e + a\,e)] + \tilde{\beta}\,\gamma|S_o| \\
&= S_o \cdot [u + d + k\,e] + (\hat{\beta} - \beta)\gamma|S_o| \\
&= S_o \cdot \left[-k\,e - \hat{\beta}\gamma\,\mathrm{sat}(S/\xi) + d + k\,e\right] + (\hat{\beta} - \beta)\gamma|S_o| \\
&= S_o \cdot \left[d - \hat{\beta}\gamma\,\mathrm{sat}(S/\xi)\right] + \hat{\beta}\gamma|S_o| - \beta\gamma|S_o| \\
&= d\,S_o - \hat{\beta}\gamma|S_o| + \hat{\beta}\gamma|S_o| - \beta\gamma|S_o| \qquad (37) \\
&\leq |d||S_o| - \beta\gamma|S| \\
&\leq |d||S_o| - (d_{max} + \eta)\gamma|S| \\
&= |d||S_o| - d_{max}\,\gamma|S_o| - \eta\,\gamma|S_o| \\
&\leq -\eta\,\gamma|S_o| \qquad (38)
\end{aligned}$$

then

$$\dot{V}(t) \leq 0 \tag{39}$$

It should be noted that in the proof the equations (28), (25), (33) and (34) and the assumptions $(\mathcal{A}2)$ and $(\mathcal{A}3)$ have been used. It has been also used that by means of S_o definition (eqn. 35), it is obtained that $S_o\,\mathrm{sat}(S/\xi) = |S_o|$.

Using the Lyapunov's direct method, since $V(t)$ is clearly positive-definite, $\dot{V}(t)$ is negative semidefinite and $V(t)$ tends to infinity as $S_o(t)$ and $\tilde{\beta}(t)$ tends to infinity, then the equilibrium at the origin $[S_o(t), \tilde{\beta}(t)] = [0, 0]$ is globally stable, and therefore the variables $S_o(t)$ and $\tilde{\beta}(t)$ are bounded. Since $S_o(t)$ is bounded then $S(t)$ is also bounded, and hence it is deduced that $e(t)$ is bounded.

From equations (25) and (29) it is obtained that

$$\dot{S}(t) \;\; = \;\; ke(t) + d(t) + u(t) \tag{40}$$

Then, from equation (40) we can conclude that $\dot{S}(t)$ is bounded because $e(t)$, $u(t)$ and $d(t)$ are bounded. Since $\dot{S}(t)$ is bounded then from equation (35) it may be deduced that $\dot{S}_o(t)$ is a bounded value.

Now, from equation (37) it is concluded that

$$\ddot{V}(t) = d\,\dot{S}_o - \beta\,\gamma\frac{d}{dt}|S_o(t)| \tag{41}$$

which is a bounded value because $\dot{S}_o(t)$ is bounded.

Under these conditions, since $\ddot{V}$ is bounded, $\dot{V}$ is a uniformly continuous function, so by means of Barbalat's lemma we can conclude that $\dot{V} \to 0$ as $t \to \infty$, which implies that $S_o(t) \to 0$ as $t \to \infty$, or equivalently that S converges to the interval $[-\xi, \xi]$ asymptotically, so under the definition of S, the error $(e = w_m - w_m^*)$ converges to a small value depending on the boundary thickness ξ.

The torque current command, $i_{sq}^*(t)$, can be obtained directly substituting eqn. (33) in eqn. (26):

$$i_{sq}^*(t) = \frac{1}{b}\left[k\,e - \hat{\beta}\gamma\,\mathrm{sat}\left(\frac{S}{\xi}\right) + a\,w_m^* + \dot{w}_m^* + f\right] \tag{42}$$

5 Simulation Results

In this section we will study the speed regulation performance of the proposed adaptive sliding-mode field oriented control under reference and load torque variations by means of simulation examples.

The block diagram of the proposed robust control scheme is presented in figure 1.

The block diagram of the proposed robust control scheme is presented in figure 1, where the block 'VSC Controller' represent the proposed adaptive sliding-mode controller, and it is implemented by equations (28), (32), and (31). The block 'limiter' limits the current applied to the motor windings so that it remains within the limit value, and it is implemented by a saturation function. The

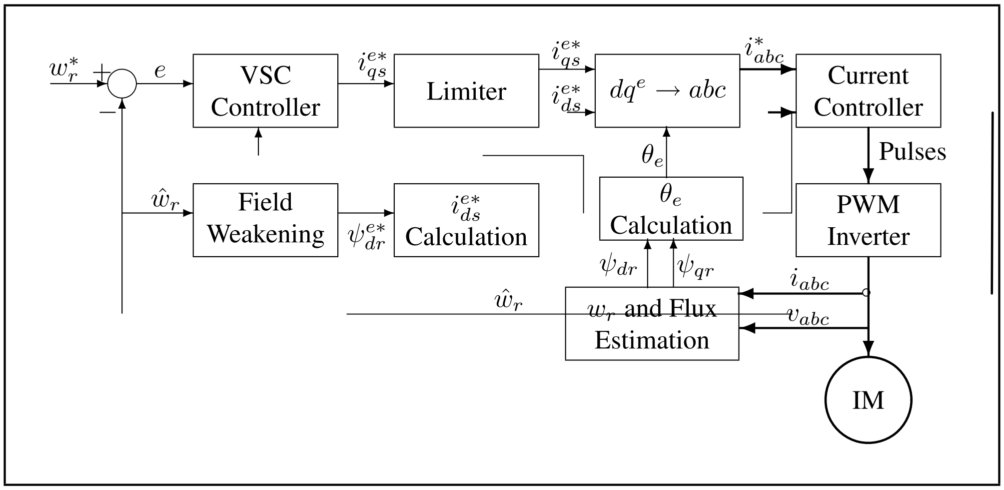

Figure 1. Block diagram of the proposed adaptive sliding-mode control

block '$dq^e \rightarrow abc$' makes the conversion between the synchronously rotating and stationary reference frames. The block 'Current Controller' consists of a three hysteresis-band current PWM control, which is basically an instantaneous feedback current control method of PWM where the actual current (i_{abc}) continually tracks the command current (i_{abc}^*) within a hysteresis band. The block 'PWM Inverter' is a six IGBT-diode bridge inverter with 780 V DC voltage source. The block 'Field Weakening' gives the flux command based on rotor speed, so that the PWM controller does not saturate. The block 'i_{ds}^{e*} Calculation' provides the current reference i_{ds}^{e*} from the rotor flux reference through the equation (20). The block 'w_r and Flux Calculation' represent the proposed rotor speed estimator and flux calculator, and is implemented by the equations (11), (1) and (2) respectively and the block 'IM' represents the induction motor.

The induction motor used in this case study is a 50 HP, 460 V, four pole, 60 Hz motor having the following parameters: $R_s = 0.087\,\Omega$, $R_r = 0.228\,\Omega$, $L_s = 35.5\,mH$, $L_r = 35.5\,mH$, and $L_m = 34.7\,mH$.

The system has the following mechanical parameters: $J = 1.662\,kg.m^2$ and $B = 0.12\,N.m.s$. It is assumed that there are an uncertainty around 20 % in the system parameters, that will be overcome by the proposed adaptive sliding control.

The following values have been chosen for the controller parameters: $k = 25$, $\gamma = 15$ and $\xi = 0.1$.

In the following examples the motor starts from a standstill state and we want the rotor speed to follow a speed command that starts from zero and accelerates until the rotor speed is $130\,rad/s$. The system starts with an initial load torque $T_L = 0\,N.m$, and at time $t = 0.6\,s$ the load torque steps from $T_L = 0\,N.m$ to $T_L = 200\,N.m$ and it is assumed that there is an uncertainty around 50 % in the load torque.

5.1 First Example

In this example, shown in Figures (2-4) it is employed the control law proposed in section 3 where it is used a signum function in the sliding mode control law. Figure 2 shows estimated rotor speed (solid line), the real rotor speed (dash-dot line) and the desired rotor speed (dashed line). As it may be observed, after a transitory time in which the sliding gain is adapted, the rotor speed tracks the desired speed in spite of system uncertainties. However, at time $t = 0.6\,s$ a little speed error can be observed. This error appears because of the torque increment at this time, and then the control system lost the so called 'sliding mode' because the actual sliding gain is too small to overcome the new uncertainty introduced in the system due to the new torque. But then, after a small time the sliding gain is adapted so that

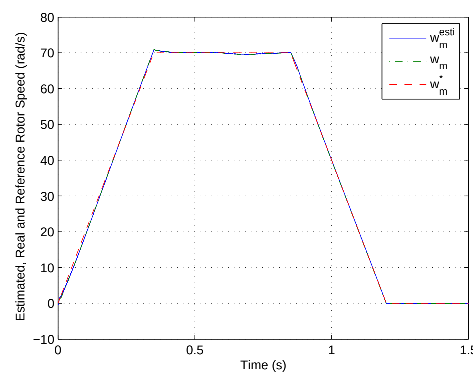

Figure 2. Estimated, Real and Reference rotor speed signals (rad/s)

this gain can compensate the system uncertainties and so the rotor speed error is eliminated. In this figure it can also be observed that the rotor speed adaptation law performs well in a low speed region. Figure 3 presents the time evolution of the estimated sliding gain. The sliding gain starts from zero and then it is increased until its value is high enough to compensate for the system uncertainties. Then at time $0.32\,s$ the sliding gain is remained constant because the system uncertainties remain constant as well. Later at time $0.6\,s$, there is an increment in the system uncertainties caused by the rise in the load torque. Therefore the sliding gain is adapted once again in order to overcome the new system uncertainties. As it can be seen in the figure, after the sliding gain is adapted it remains constant again, since the system uncertainties remains constant as well.

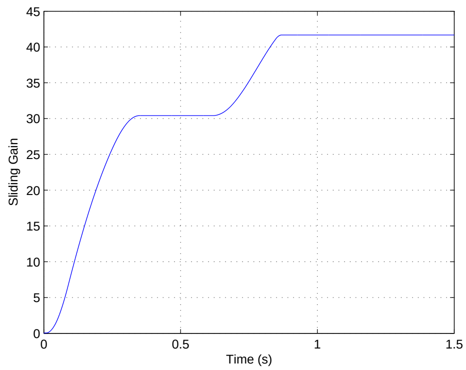

Figure 3. Estimated sliding Gain

It should be noted that the adaptive sliding gain allows to employ a smaller sliding gain. In this way it is not necessary to choose the slading gain value high enough to compensate all the possible system uncertainties as used in conventional sliding control laws. With the proposed adaptive scheme the sliding gain is adapted (if necessary) when a new uncertainty appears in the system in order to surmount this uncertainty. Figure 4 shows the motor torque.

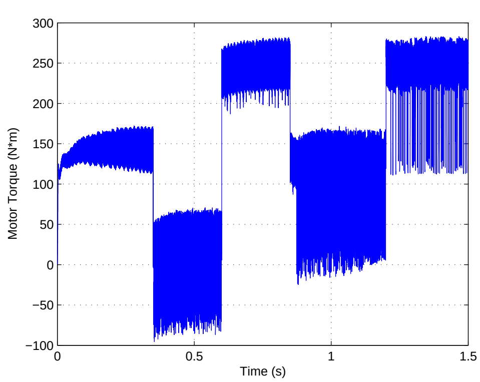

Figure 4. Motor torque (N.m)

This figure shows that in the initial state, the motor torque has a high initial value in the speed acceleration zone because it is necessary a high torque to increment the rotor speed owing to the rotor inertia, then the value decreases in a constant region and finally increases due to the load torque increment. In this figure it may be observed that in the motor torque appears the so-called chattering phenomenon due to the signum function presented in the control law. It should be noted that the chattering involves high control activity and may further excite high-frequency dynamics. This undesirable effect can be avoided using the modified adaptive sliding control law proposed in Section 4, as it is shown in the next simulation results.

5.2 Second Example

In this second example, shown in Figures (5-7) it is employed the control law proposed in section 4 where the control law is smoothed out within a boundary layer in order to avoid the chattering phenomenon that is undesirable in practice.

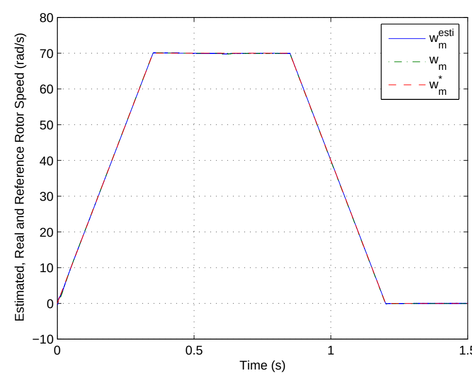

Figure 5. Estimated, Real and Reference rotor speed signals (rad/s)

Figure 5 shows estimated rotor speed (solid line), the real rotor speed (dash-dot line) and the desired rotor speed (dashed line). Similarly to the previous example, after a transitory time in which the sliding gain is adapted, the rotor speed tracks the desired speed in spite of system uncertainties. Figure 6 presents the time evolution of the estimated sliding gain. As before, the sliding gain starts from zero and then it is increased until its value is high enough in order to compensate the system uncertainties. Then at time $t = 0.6\,s$, the sliding gain is adapted once again in order to overcome the new system uncertainties caused by the the rise in the load torque. Figure 7 shows the motor torque. As in the first example the motor torque presents a high initial value in the speed acceleration zone, then the value decreases in a constant region and at time $t = 0.6\,s$ the motor torque increases due to the load torque increment. However, as it may be observed, unlike the previous example, in the present example the chattering phenomenon

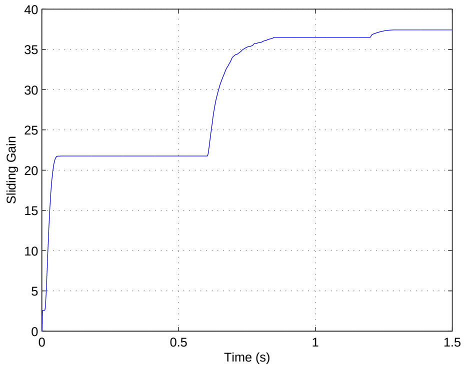

Figure 6. Estimated sliding Gain

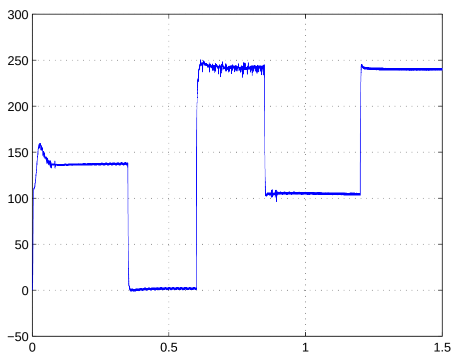

Figure 7. Motor torque (N.m)

does not appear in the motor torque, because the control signal is smoothed out within a boundary layer by means of a saturation function instead of signum function. In consequence, the smoothed control law proposed in Section 4 presents lower control activity than the previous one presented in Section 3, and therefore the motion of the motor drive in this case would be softer than in the previous case.

6 Conclusion

In this paper a sensorless adaptive sliding mode vector control has been presented. The rotor speed adaptation law is based on stator current equations and rotor flux equations in the stationary reference frame, and using simulation examples it is demonstrated that this adaptation law performs well in a low speed region. It is proposed a new adaptive variable structure control that is robust under uncertainties caused by parameter error or by changes in the load torque. Moreover, the proposed variable structure control incorporates an adaptive algorithm to calculate the sliding gain value. The adaptation of the sliding gain, on the one hand avoids the necessity of computing the upper bound of the system uncertainties, and on the other hand allows to employ as smaller sliding gain as possible in order to overcome the actual system uncertainties. Therefore, the control signal of our proposed variable structure control schemes will be smaller that the control signals of the traditional variable structure control schemes. Finally, by means of simulation examples, it has been shown that the proposed control scheme performs reasonably well in practice, and that the speed tracking objective is achieved under uncertainties in the parameters and load torque.

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