SUB-OPTIMAL ALGORITHM SECOND-ORDER SLIDING MODE CONTROL FOR A SYNCHRONOUS RELUCTANCE MOTOR SPEED DRIVE

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ABSTRACT
A sub-optimal algorithm second-order sliding mode controller (SOSMC) was presented for a synchronous reluctance motor (SynRM) speed drive. SOSMC is an effective tool for the control of uncertain nonlinear systems since it overcomes the main chattering drawback of conventional sliding mode control. The practical implementation of SOSMC has simple control laws and assures an improvement in sliding accuracy with respect to conventional sliding mode control. This paper proposes a control scheme based on sub-optimal algorithm SOSMC. The proposed SOSMC is robust against motor parameter variations and mitigates chattering phenomenon. Experiments were conducted to validate the proposed method.

Keywords: sub-optimal algorithm; second order sliding mode control; synchronous reluctance motor; chattering phenomenon.

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UN ALGORITHME SOUS-OPTIMAL PAR MODE GLISSANT DU SECOND ORDRE POUR LA COMMANDE DE VITESSE D’UN MOTEUR À RELUCTANCE SYNCHRONE

RÉSUMÉ
Un algorithme sous-optimal par mode glissant de second ordre est présenté pour la commande d’un moteur à reluctance synchrone (SOSMC). Ce dernier est un outil efficace pour la commande de systèmes non-linéaires incertains parce qu’il élimine l’inconvénient du cliquettage de la commande par mode glissant conventionnel. La réalisation pratique du SOSMC comporte des règles de commande simples, et garantit une amélioration de la précision du glissement par rapport à la commande par mode glissant conventionnel. Cet article propose un système de commande basé sur un algorithme sous-optimal SOSMC. Cet algorithme est robuste envers les paramètres variables et atténue le phénomène du cliquettage. Des expériences ont été menées pour valider la méthode proposée.

Mots-clés: algorithme sous-optimal; commande par mode glissant de second ordre; moteur à reluctance synchrone; phénomène de cliquettage.
NOMENCLATURE

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
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<tbody>
<tr>
<td>$V_{ds}$, $V_{qs}$</td>
<td>direct axis and quadrature axis terminal voltages (V)</td>
</tr>
<tr>
<td>$i_{ds}$, $i_{qs}$</td>
<td>direct axis and quadrature axis terminal currents (A)</td>
</tr>
<tr>
<td>$L_{ds}$, $L_{qs}$</td>
<td>direct axis and quadrature axis magnetizing inductances (H)</td>
</tr>
<tr>
<td>$R_s$</td>
<td>stator resistance (Ω)</td>
</tr>
<tr>
<td>$\omega_r$</td>
<td>rotor speed</td>
</tr>
<tr>
<td>$P_{ole}$</td>
<td>numbers of poles</td>
</tr>
<tr>
<td>$T_e$</td>
<td>electromagnetic torque (N·m)</td>
</tr>
<tr>
<td>$T_L$</td>
<td>load torque (N·m)</td>
</tr>
<tr>
<td>$J_m$</td>
<td>rotor inertia moment (Kg·m²)</td>
</tr>
<tr>
<td>$B_m$</td>
<td>viscous friction coefficient (Nt·m/rad/sec)</td>
</tr>
<tr>
<td>$I_s$</td>
<td>peak value of motor input phase current</td>
</tr>
<tr>
<td>$e(t)$</td>
<td>velocity error</td>
</tr>
<tr>
<td>$\omega_{ref}$</td>
<td>velocity command</td>
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<tr>
<td>$S$</td>
<td>sliding function</td>
</tr>
<tr>
<td>$u_{eq}(t)$</td>
<td>equivalent control input</td>
</tr>
<tr>
<td>$u_n(t)$</td>
<td>uncertainty switching control input</td>
</tr>
<tr>
<td>$v(t)$</td>
<td>scalar control input</td>
</tr>
<tr>
<td>$y_1(t)$</td>
<td>sliding function</td>
</tr>
<tr>
<td>$y_2(t)$</td>
<td>unmeasurable variable but with a possibly known sign</td>
</tr>
<tr>
<td>$\hat{y}_{1M}$</td>
<td>the last value of the $y_1(t)$ function</td>
</tr>
<tr>
<td>$\alpha^*$</td>
<td>modulation factor</td>
</tr>
<tr>
<td>$\beta$</td>
<td>anticipation factor</td>
</tr>
<tr>
<td>$V_M$</td>
<td>magnitude factor</td>
</tr>
</tbody>
</table>

1. INTRODUCTION

In the past decades, DC motors have been widely used in factory automation as high-performance drives. However, the mechanical commutators and brush assembly make dc motors much more expensive than ac motors. Furthermore, the use of mechanical commutators may generate undesired sparks, which are not allowed in some applications. The instinctive drawbacks of dc drives have reminded incessant attempts to find better solutions instead of dc drives. An attempt was made to use AC drives including synchronous and induction drives. Nevertheless, the synchronous motor needs slip rings, brushes, and insulated winding on the rotor. On the other hand, the induction motor has not been widely used in high-performance drives due to its high nonlinearity and time-varying characteristics [1–9].

A fast error-free dynamic response is offer the primary concern in control systems. Practical servo systems have parameter variations and external load disturbances. In order to overcome uncertainty, sliding mode control (SMC) [7–9] was developed. Variable structure control (VSC) or SMC has been known as a very effective way for the position and velocity control of motors because it possess many advantages, such as insensitivity to parameter variations and external disturbance rejection, and fast dynamic responses. Therefore, VSC has been widely used in dc and ac motor drives [7–11]. The system dynamics of a VSC system can be divided into two modes: the reaching mode and sliding mode. The robustness of a VSC system inheres in its sliding mode, but not in its reaching mode. In other words, the system performance of the VSC system is mainly designed in the sliding mode. Therefore, the reaching mode is designed to force the system trajectories into the sliding phase as soon as possible. SMC is an effective and robust technology for parameter variation and external disturbance rejection. SMC is a robust for nonlinear systems. Discontinuous systems require an infinite switching frequency. Therefore, reducing chattering is very important for SMC. Second-order SMC (SOSMC) [12, 13] is robust against model uncertainties and external disturbances, while mitigating chattering. However, few studies have been conducted on SOSMC for SynRMs.
Unlike conventional first-order SMC, SOSMC belongs to higher-order sliding mode (HOSM). Levant [10] determined the relationship between accuracy and sampling time for HOSM. Bartolini et al. [11] developed a HOSM model by the classical bang-bang optimal control strategy. The proposed algorithm does not require the use of observers and differential inequalities and solve the problem of chattering elimination in VSC. They found that HOSM has simple construction and is robust against system structure variability. Bartolini et al. [13, 14] proposed another HOSM scheme (Simplex sliding mode control) for the multi-input nonlinear uncertain systems, which only partitions the state space into m+1 regions. Clearly, the number of regions is decreased and as a result the chattering problem can be improved.

In application parts, Damiano et al. [15] adopted a novel scheme based on SOSMC technique for a permanent magnet (PM) DC motor speed control. The proposed scheme shows the robustness against heavy motor parameters variations and the avoidance of the current loop implementation. Ferrara and Rubagotti [16] applied SOSMC for a current-fed induction motor and verified by simulation results. Levant proved finite time convergence using point-to-point method for state trajectories [17, 18]. Lyapunov function was found and analyzed in [19, 20], Lyapunov function with finite convergence time as a solution to partial differential equation was offered in [21, 22]. Utkin [23] performed the analysis in time domain directly. Zhao et al. [24] presents a speed estimation scheme based on second-order sliding-mode supertwisting algorithm (STA) and model reference adaptive system (MRAS) estimation theory for a sensorless induction motor drives. Ding et al. [25] develops a two-sliding mode control algorithm for a class of non-linear systems under the assumption that the uncertainty of the sliding mode dynamics can be bounded by a positive function. Lin et al. [26] proposes an intelligent SOSMC using a wavelet fuzzy neural network with an asymmetric membership function estimator to control a six-phase permanent magnet synchronous motor for an electric power steering system. Chang [27] presents a dynamic compensator-based second-order sliding mode control algorithm without using any observer structure to estimate the velocity for an Inverted pendulum mechanical system.

Only five studies [28–32] have applied SOSMC to SynRM control. [28] adopted super-twisting sliding mode controller as a reduced order observer of the rotor fluxes estimation for a synchronous motors in numerical simulation. Mohamadian et al. [29] used sub-optimal SOSMC. Chiang et al. [30] used sub-optimal SOSMC in simulations. Boroujeni et al. [31, 32] adopted super-twisting SOSMC in simulations. Hence, this paper proposes a control scheme based on sub-optimal SOSMC that is verified by experiments.

The rest of this paper is organized as follows. SynRM modeling in the synchronously rotating rotor reference frame is discussed in Section 2. In Section 3, the vector control of SynRM is introduced. In Section 4, the integral variable structure speed controller is described. In Section 5, the sub-optimal algorithm SOSMC is derived. The proposed speed controller is implemented using a PC-based SynRM drive. In Section 6, experimental results show that the proposed sub-optimal algorithm SOSMC controller provides high-performance dynamic characteristics and robustness against parameter variation and external load disturbances. Finally, conclusions are presented in Section 7.

2. SYNRM MODELING

For analysis, the three-phase fixed a-b-c frame of reference in the stator can be converted into a synchronously rotating rotor reference frame using Park’s transformation. The d-q equivalent circuit of the ideal SynRM model is shown in Fig. 1.

\[ V_{ds} = R_s i_{ds} + L_s \frac{di_{ds}}{dt} - \omega_r L_{qs} i_{qs} \]  \hspace{1cm} (1)

\[ V_{qs} = R_s i_{qs} + L_{qs} \frac{di_{qs}}{dt} + \omega_r L_{ds} i_{ds} \]  \hspace{1cm} (2)
The corresponding electromagnetic torque $T_e$ is

$$ T_e = \frac{3}{4} P_{\text{ole}} (L_{ds} - L_{qs}) i_{ds} i_{qs} \tag{3} $$

The corresponding motor dynamic equation is

$$ T_e - T_L = J_m \frac{d}{dt} \omega_r + B_m \omega_r \tag{4} $$

where $V_{ds}$ and $V_{qs}$ are the direct axis (d axis) and quadrature axis (q axis) terminal voltages, respectively; $i_{ds}$ and $i_{qs}$ are, respectively, the direct axis and quadrature axis terminal currents or the torque producing current; $L_{ds}$ and $L_{qs}$ are the direct axis and quadrature axis magnetizing inductances, respectively; $R_s$ is the stator resistance; and $\omega_r$ is the speed of the rotor. $P_{\text{ole}}$, $T_L$, $J_m$, and $B_m$ are the poles, the torque load, the inertia moment of the rotor, and the viscous friction coefficient, respectively.

3. SYNRM VECTOR CONTROL

Vector control utilizes the transformation method of the reference frame. It can transform the $a$-$b$-$c$ axis fixed reference frame into the $d$-$q$ axis synchronously rotating reference frame. For an AC motor, the output torque of a SynRM can be adjusted by controlling the currents of the axis and axis appropriately. By adjusting current angle, Betz et al. [5] offered a comprehensive approach about SynRM control strategy. The inherent disadvantage of SynRM is minor average torque. To have the property of maximum torque per ampere generation, the maximum torque control (MTC) strategy [6] will be adopted in this study. Let the current angle $\phi = \tan^{-1}(i_{qs}/i_{ds})$, the peak value of motor input phase current is $I_s = \sqrt{i_{ds}^2 + i_{qs}^2}$. The electromagnetic torque equation (3) can be rewritten as

$$ T_e = \frac{3}{2} \frac{P_{\text{ole}} I_s^2}{2} (L_{ds} - L_{qs}) \sin 2\phi $$

The maximum torque current angle $\phi$ for the MTC strategy is $\phi = \pm \pi/4$. The torque current commands are the following:

$$ i_{ds}^* = \sqrt{\frac{|T_e|}{\frac{3}{2} P_{\text{ole}} (L_{ds} - L_{qs})}} \cos \left(\frac{\pi}{4}\right), \quad i_{qs}^* = \text{sign}(T_e) \sqrt{\frac{|T_e|}{\frac{3}{2} P_{\text{ole}} (L_{ds} - L_{qs})}} \sin \left(\frac{\pi}{4}\right) \tag{5} $$

Fig. 1. $d$-$q$ axis equivalent circuit of a SynRM.
4. INTEGRAL VARIABLE STRUCTURE SLIDING MODE CONTROLLER

The motor dynamics equation (4) can be rewritten as

\[ \dot{\omega}_r = \left( -\frac{B_m}{J_m} \right) \omega_r + \frac{1}{J_m} (T_e - T_L) = a \omega_r + b (T_e - T_L) = (a_0 + \Delta a) \omega_r + (b_0 + \Delta b)(T_e - T_L) \]

\[ = a_0 \omega_r + b_0 (u(t) + f) \quad (6) \]

where

\[ a \equiv -\frac{B_m}{J_m} = a_0 + \Delta a, \quad b \equiv \frac{1}{J_m} = b_0 + \Delta b, \quad u \equiv T_e \]

\[ f \equiv \frac{1}{b_o} (\Delta a \omega_r + \Delta b u(t) - b T_L) \]

\[ J_m \equiv J_o + \Delta J, \quad B_m \equiv B_o + \Delta B \]

The subscript index \( o \) indicates the nominal system value; \( \Delta \) represents uncertainty, and \( f \) represents the lumped uncertainties.

Define the velocity error as \( e(t) = \omega_{\text{ref}} - \omega_r \), where \( \omega_{\text{ref}} \) is the velocity command. The velocity error differential equation of a SynRM can be expressed as

\[ \frac{de(t)}{dt} = \frac{d\omega_{\text{ref}}}{dt} - \frac{d\omega_r}{dt} = \dot{\omega}_{\text{ref}} - a_0 \omega_r - b_0 [u(t) + f] \quad (7) \]

The sliding function \( S \) is combined with the integration of the error as

\[ S = e(t) = c \int_{-\infty}^{t} e(\tau) d\tau, \quad c > 0 \quad (8) \]

The input control \( u(t) \) (the electromagnetic torque \( T_e \)) can be defined as

\[ u(t) = u_{\text{eq}}(t) + u_n(t) \quad (9) \]

where \( u_{\text{eq}}(t) \) is used to control the overall behavior of the system and \( u_n(t) \) is used to suppress parameter uncertainties and to reject disturbances. To satisfy the equivalent control concept \( \dot{S} = \dot{e} = 0 \), we obtain

\[ u_{\text{eq}} = \frac{1}{b_o} (\dot{\omega}_{\text{ref}} - a_0 \omega_r + ce) \quad (10) \]

To satisfy the reaching condition \( S(t) \dot{S}(t) \leq 0 \), we have

\[ S \dot{S} = -S[b_0(u_n + f)] \quad (11) \]

Let \( |f| \leq K \), \( \eta \) is a positive constant. The uncertain nonlinear switch control input can be defined as

\[ u_n = \left( K + \frac{\eta}{b_o} \right) \text{sign}(S) \quad (12) \]
5. SUB-OPTIMAL ALGORITHM SECOND-ORDER SLIDING MODE CONTROLLER

In conventional SMC design, the control target is to move the system state into sliding surfaces \( S = 0 \). In HOSM of the \( r \)th order sliding mode, it aims for \( S = \dot{S} = \cdots = S^{(r-1)} = 0 \). System (1) evolves featuring a SOSMC mode on the sliding manifold (2). If and only if its state trajectories lie on the intersection of the two manifolds \( S = 0 \) and \( \dot{S} = 0 \) in the state space.

The sub-optimal algorithm has been developed from the time-optimal control rules [12]. The state trajectory of the \( S \) and \( \dot{S} \) phase plane is shown in Fig. 2. Finally, it converges to the origin of the phase plane.

Consider sliding variable dynamics given by a system with a relative degree of two:

\[
\begin{align*}
\dot{y}_1 &= y_2 \\
\dot{y}_2 &= \varphi(y(t), t) + \gamma(y(t), t)v(t)
\end{align*}
\]

(13)

where \( y_1(t) \) is the sliding function \( S \), \( v(t) \) is the scalar control input, \( y_2(t) \) is an unmeasurable variable but with a possibly known sign, and \( \varphi(y(t), t) \) and \( \gamma(y(t), t) \) are uncertain functions such that

\[
|\varphi(y(t), t)| \leq \Phi, \quad 0 < \Gamma_m \leq \gamma(y(t), t) \leq \Gamma_M, \quad 3\Gamma_m > \Gamma_M
\]

(14)

State equation (14) can be given uncertain dynamic response to satisfy Eq. (15). Then, we adopt Eq. (16) digital feedback controller, so the system state can converges to the origin nearby of the phase plane [10, 12, 15, 29]:

\[
\begin{align*}
\dot{y}_{1M} &= y_1[-1] = y_1(0), \quad y[-2] = 0, \quad \Delta[k] = [y_1[k-1] - y_1[k]]y_1[k] \\
\dot{y}_{1M}[k] &= \begin{cases} 
\dot{y}_{1M}[k-1], & \text{if } \Delta[k] > 0 \\
y_1[k-1], & \text{if } \Delta[k] \leq 0
\end{cases} \quad k = 0, 1, 2, \ldots
\end{align*}
\]

\[
\alpha = \begin{cases} 
\alpha^*, & \text{if } [y_1[k] - \beta\dot{y}_{1M}][\dot{y}_{1M} - y_1[k]] > 0 \\
1, & \text{if } [y_1[k] - \beta\dot{y}_{1M}][\dot{y}_{1M} - y_1[k]] \leq 0
\end{cases}
\]

(15)

\[
V_M = \frac{4\Phi}{3\Gamma_m - \alpha^*\Gamma_M} \left[ 1 + \sqrt{1 + \frac{3\Gamma_m - \alpha^*\Gamma_M}{4\Gamma_M}} \right], \quad v(t) = -\alpha V_M \text{sign}[y_1[k] - \beta\dot{y}_{1M}]
\]
where $\hat{y}_{1M}$ represents the last value of the $y_1(t)$ function. In practical cases $\hat{y}_{1M}$ can be estimated by checking the sign of the quantity $\Delta[k] = [y_1[k-1] - y_1[k]]y_1[k]$. $\alpha^*$ is the modulation factor, $\beta$ is the anticipation factor and $V_M$ is a positive coefficient called magnitude factor. By properly setting the controller parameters $\alpha^*$, $\beta$, and $V_M$, it turns out that both $S$ and $\dot{S}$ converge to zero in finite time [33]. Equation (4) can be rewritten as

$$\frac{d\omega_r}{dt} = \frac{1}{J_m} (T_e - T_L - B_m \omega_r)$$

(16)

The state variable is defined as

$$\begin{cases}
  x_1(t) = \int_{-\infty}^{t} x_2(\tau)d\tau \\
  x_2(t) = e(t) = \omega_{\text{ref}} - \omega_r
\end{cases}$$

(17)

The initial value of integration $x_1(t_0)$ can be expressed as

$$x_1(t_0) = \int_{-\infty}^{t_0} x_2(\tau)d\tau = \frac{\omega_r - \omega_{\text{ref}}}{c}$$

(18)

Assume $\dot{\omega}_{\text{ref}} = 0$, $\omega_{\text{ref}}$ is a constant or slowing time varying command, then the system state equation of a SynRM can be expressed as

$$\begin{cases}
  \dot{x}_1 = x_2 \\
  \dot{x}_2 = -\frac{B_m}{J_m} x_2 + \frac{1}{J_m} T_L - \frac{1}{J_m} T_e + \frac{B_m}{J_m} \omega_{\text{ref}} = -\frac{B_m}{J_m} x_2 + \frac{1}{J_m} T_L + \hat{u}(t)
\end{cases}$$

(19)

where

$$\hat{u}(t) = -\frac{1}{J_m} T_e + \frac{B_m}{J_m} \omega_{\text{ref}}$$

(20)

The sliding functions $y_1$ and $y_2$ are defined as

$$\begin{cases}
  y_1 = x_2 + cx_1 \\
  y_2 = \dot{y}_1
\end{cases}$$

(21)

Then, the state equation of uncertain second-order system can be expressed as

$$\begin{cases}
  \dot{y}_1 = y_2 \\
  \dot{y}_2 = \left(-\frac{B_m}{J_m} + c\right) x_2 + \frac{1}{J_m} T_L + v(t)
\end{cases}$$

(22)

It satisfies the conditions

$$\begin{cases}
  B_m \leq B_{\text{max}} \\
  J_{\text{min}} \leq J_m \leq J_{\text{max}} \\
  |\dot{\omega}_r| \leq \dot{\omega}_{\text{max}} \\
  |\dot{T}_L| \leq \dot{T}_{L_{\text{max}}}
\end{cases}$$

(23)

The system control signal input $v(t)$ can be expressed as

$$v(t) = -\frac{1}{J_m} \dot{T}_e(t)$$

(24)

According to Eq. (24), the practical controllable signal $T_e$ of a SynRM is a continuous controllable signal that integrates the defined input controllable signal from Eq. (15), which can mitigates the chattering.
Fig. 3. (a) Sub-optimal SOSMC speed control block diagram of SynRM servo drive. (b) The proposed hardware architecture and the photo of experimental equipment.

6. EXPERIMENTAL RESULTS

Block diagram of the proposed speed control system is shown in Fig. 3(a). The proposed hardware architecture and the photo of experimental equipment are shown in Fig. 3(b).

This system has a hardware drive circuit, a SynRM, mechanical loads, and auxiliary circuitry for control and measurement. In Fig. 3(a), a vector control algorithm is implemented. Reference speed is compared
Fig. 4. Experimental results of SMC due to $\omega_{ref} = 600$ rpm under no machine load in nominal case of motor inertia and friction coefficient. (a) Rotor velocity response, (b) control signal response.

Fig. 5. Experimental results of sub-optimal algorithm SOSMC for $\omega_{ref} = 600$ rpm under no machine load in the nominal case of motor inertia and friction coefficient. (a) Rotor velocity response, (b) control signal response.

with actual speed, measured from an absolute encoder. The speed error and rotor position are the inputs to the SOSMC block. The command torque signal is the output of the controller. The proposed system has verified the global behavior and the robustness to parameter variations and load disturbances. We simulate it by using the Simulink of Matlab. The experimental system adopts the dSPACE DS1104 control board to act as the digital speed control system. The SynRM is driven using IGBTs by a three-phase voltage space vector pulsewidth modulation (VSVPWM) inverter. It is shown that the IGBT switch status may not change in 120$^\circ$ sectors for each phase in a cycle, and the positions of the non-switching sectors can be adjusted continuously by allocating the duty time of the two zero vectors properly.

The controller was implemented using a DS1104 controller board (dSPACE, Inc., Germany) with a fixed point DSP TMS320F240. DS1104 is designed for a standard PC environment. The synchronous reluctance motor modeled in this paper is a 0.37-kW, 2-pole, 230-V, 4.7-A, 60-Hz, 3600-rpm machine. The machine parameters are as follows: (1) stator resistance $R_S = 4.2 \Omega$, (2) direct axis magnetizing inductance $L_{ds} = 328$ mH, (3) quadrature axis magnetizing inductance $L_{qs} = 181$ mH, (4) rotor inertia $J = 0.00076$ Kg-m$^2$, (5) friction coefficient $B_m = 0.00012$ Nt-m/rad/sec, (6) sliding function $c = 6$, (7) sampling time 0.0003 sec, (8) modulation factor $\alpha^* = 0.5$, and (9) anticipation factor $\beta = 0.9$. 

Fig. 6. Experimental results of SMC for $\omega_{\text{ref}} = 600$ rpm under a 0.3 Nt-m machine load at the beginning and a 0.9 Nt-m machine load at 5 seconds in the nominal case of motor inertia and friction coefficient. (a) Rotor velocity response, (b) control signal response.

Fig. 7. Experimental results of sub-optimal algorithm SOSMC for $\omega_{\text{ref}} = 600$ rpm under a 0.3 Nt-m machine load at the beginning and a 0.9 Nt-m machine load at 5 seconds in the nominal case of motor inertia and friction coefficient. (a) Rotor velocity response, (b) control signal response.

Figure 4 is the experimental responses of SMC velocity and control signal for command $\omega_{\text{ref}} = 600$ rpm under no machine load in nominal case motor inertia and friction coefficient.

Figure 5 is the experimental responses of SOSMC velocity and control signal for command $\omega_{\text{ref}} = 600$ rpm under no machine load in nominal case motor inertia and friction coefficient.

From Figs. 4 and 5, the proposed SOSMC mitigates the chattering drawback of SMC apparently. Figure 6 is the experimental responses of SMC velocity and control signal for $\omega_{\text{ref}} = 600$ rpm under a 0.3 Nt-m machine load at the beginning and a 0.9 Nt-m machine load at 5 seconds in the nominal case motor inertia and friction coefficient. Figure 7 is the experimental responses of SOSMC velocity and control signal for $\omega_{\text{ref}} = 600$ rpm under a 0.3 Nt-m machine load at the beginning and a 0.9 Nt-m machine load at 5 seconds in the nominal case motor inertia and friction coefficient. From Figs. 4 to 7, the proposed SOSMC with sliding phase layer concept is robust against motor parameter variation and external disturbances. The needing sliding phase layer of the proposed SOSMC is thinner than the conventional SMC. Thus, it can reduce the steady state error and improve the accuracy of SMC.
7. CONCLUSIONS

A sub-optimal algorithm SOSMC design for robust stabilization and disturbance rejection of a SynRM drive was proposed. The experimental results show good performance for SOSMC under uncertain load subject to variations in inertia and system friction. Comparing with the SMC controller under a no-load condition, the proposed SOSMC controller can mitigate chattering phenomenon. However, it has a sinusoidal oscillation situation occurring in the steady state due to control signal converting. It also has overshoot occurring in the initial period during loading disturbance. These circumstances can be improved by properly setting the controller parameters $\alpha^*, \beta$, and $V_M$. Thus, it can increase the control signal to solve the above problem and have better responses in the follow-up loading period. The real implementation of proposed SOSMC implies very simple control laws and assures an improvement of the sliding accuracy with respect to real 1-sliding mode control. The derived SOSMC laws are continuous, and thus eliminate the chattering effect. The proposed SOSMC provides a faster and better response under parameter variation and external disturbances.

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