2011 American Control Conference
on O'Farrell Street, San Francisco, CA, USA
June 29 - July 01, 2011

# Passivity-Based Adaptive Sliding-Mode Speed Control for IPMSM Drive Systems

Jen-te Yu, Cheng-Kai Lin, Li-Chen Fu*, and Tian-Hua Liu

***Abstract*-A passivity-based adaptive sliding-mode control is proposed for speed tracking of interior permanent magnet synchronous motor drive systems. Firstly, a nonlinear model of the IPMSM is given with uncertainties embeded. Through adaptive feedback passivation design, the closed-loop system is shown to be feedback equivalent to a strictly passive system with a designated input. The unknown system parameters are dealt with by designed adaptation laws in parallel with the design of the controller. Maximum torque per ampere condition is met through the design of d- and q-axis currents, which serve as the inputs to the motor. Asymptotic stability of closed loop system is proven by passivity theorem and Barbalat's lemma. Simulation results show good speed tracking response and good performance.**

## I. INTRODUCTION

The interior permanent magnet synchronous motor (IPMSM) has been widely used in industry because of its high efficiency, high torque/ampere ratio, and rugged structure. Several passivity-based controllers have been developed for AC and DC motors to enhance their performance in the past [1]-[3]. Gökdere and Simaan proposed a passivity-based method for induction motor control [1]. The performance of the method still relies on two proportional-integral (PI) controllers to achieve the current-loop control. Although stability of the system can be guaranteed by passivity theorem, poor control performance caused by parametric uncertainties might arise. A passivity-based composite adaptive position control scheme for an induction motor was attempted in [2]. This method can compensate some unknown but slowly varying parametric uncertainties in the system. Another passivity-based sliding-mode controller was proposed to improve the performance of induction motors in [3]. The external load torque was estimated by an adaptive law. However, signum function is used in the controller, resulting in chattering.

Through the feedback passivation approach, the input-output feedback linearization design method can make the closed-loop system strictly passive with respect to the designated input [4]. Following that, the asymptotic stability of the closed-loop system can be guaranteed, based on passivity theory. Adaptive backstepping sliding-mode control methodologies are widely used for motion control of AC motors. For example, Lin *et al* proposed an adaptive backstepping sliding-mode control for induction motor drives in [5]. Using that technique, one can effectively reduce the effect of the unknown but slowly varying parameters on the system. Although good performance can be obtained, the systems may still suffer chattering due to switching. To overcome the chattering problem resulting from traditional sliding-mode control approach, several papers have addressed the issue [6], [7]. For example, Xu proposed a chattering free robust control for nonlinear systems [6]. The time-varying feedback gain, however, is very difficult to obtain in real-time implementation. Shahnazi *et al.* proposed yet another novel adaptive fuzzy PI sliding-mode control for induction and DC servomotors in [7]. Through simulation results, their method showed good tracking performance without chattering.

We propose in this paper a passivity-based adaptive sliding-mode controller for speed tracking of the IPMSM. To design the controller that satisfies the passivity property, we employ an unconventional initial step in this paper in which the time-derivative of the storage function is provided firstly to start the design procedure but only symbolically. The specifications of the partial-derivatives of the storage function's components are tailored so as to fulfill the energy balance equation. After that, an adaptive sliding-mode control is used to enhance the performance against system uncertainties and external load torque disturbance. In addition, to reduce chattering caused by switching functions, a PI-saturation function is utilized to approximate the signum function within the boundary layer. Different from traditional passivity-based approaches toward this problem which keep the d-axis current constant, our method can maintain the d- and q-axis currents on the MTPA trajectory. To the authors' best knowledge, no researchers have published papers of passivity-based adaptive sliding-mode MTPA control for IPMSM drive systems. This motivates us to study the issue. Our simulation results do show feasibility of the proposed control scheme.

The paper is organized as follows. An introduction is given firstly. In Section II, a mathematical model of IPMSM is provided. The design of passivity-based adaptive sliding-mode MTPA control is presented in Section III. Simulation results are shown and discussed in Section IV. Section V concludes this paper.

This work was supported by National Science Council, Taiwan under the grant NSC 97-2221-E-002-178-MY3.

J. Yu and C. K. Lin are with the Department of Electrical Engineering, National Taiwan University, Taipei, Taiwan.

L. C. Fu is with the Department of Electrical Engineering and Department of Computer Science and Information Engineering, National Taiwan University, Taipei, Taiwan (e-mail: lichen@ntu.edu.tw).

T. H. Liu is with the Department of Electrical Engineering, National Taiwan University of Science and Technology, Taipei, Taiwan.

## II. MODELING OF IPMSM

The governing (voltage) equations in d-q axis for the IPMSM can be expressed as:

$$v_d = r_s i_d + L_d \frac{di_d}{dt} - P_o \omega_{rm} L_q i_q \quad (1)$$

$$v_q = r_s i_q + L_q \frac{di_q}{dt} + P_o \omega_{rm} L_d i_d + P_o \omega_{rm} \lambda_m \quad (2)$$

where $v_d$ and $v_q$ are d- and q-axis voltages, $r_s$ is stator resistance, $i_d$ and $i_q$ are d- and q-axis currents, $L_d$ and $L_q$ are d- and q-axis inductances, d/dt is the differential operator, $P_o$ is the number of pole-pairs, and $\omega_{rm}$ is motor speed. The equation for the speed is

$$\frac{d\omega_{rm}}{dt} = \frac{1}{J_m}(T_e - T_L - B_m \omega_{rm}) \quad (3)$$

in which

$$T_e = \frac{3}{2} P_o \left[ \lambda_m i_q + (L_d - L_q) i_d i_q \right] \quad (4)$$

Here $T_e$ is electromagnetic torque of the motor, $T_L$ is external load torque, $J_m$ is moment of inertia of rotor, $B_m$ is viscous friction coefficient, and $\lambda_m$ is flux linkage.

## III. PASSIVITY-BASED ADAPTIVE SLIDING-MODE MTPA CONTROL

According to (3), the dynamic equation of speed with system uncertainties can be expressed as [5]:

$$\dot{\omega}_{rm} = \frac{T_e}{J_{mo}} - \frac{\omega_{rm} B_m}{J_{mo}} - \frac{T_L}{J_{mo}} + \frac{F_3}{J_{mo}} \quad (5)$$

where

$$|F_3| < \bar{F}_3 \quad (6)$$

Here $F_3$ stands for lumped uncertainty whose upper bounded is denoted as $\bar{F}_3$, which is known. $J_{mo}$ refers to the nominal value of $J_m$. We will assume the lumped uncertainty $F_3$ is constant (but unknown) within a sampling period due to the fact that its variation is negligible in a sampling interval. As a result, the lumped uncertainty $F_3$ can be roughly estimated by an adaptive estimation law [5]. Due to their very different natures, the load torque and $F_3$ are not aggregated into one entity.

### A. *Feedback Passivation*

Consider the dynamic equation (3), and define the speed tracking error as follows, where and throughout this paper a variable with a superscript "*" refers to its desired value.

$$e_1 = \omega_{rm}^* - \omega_{rm} \quad (7)$$

The time derivative of speed tracking error is written as

$$\begin{aligned} \dot{e}_1 &= \dot{\omega}_{rm}^* - \dot{\omega}_{rm} \\ &= \dot{\omega}_{rm}^* - \frac{T_e}{J_m} + \frac{\omega_{rm} B_m}{J_m} + \frac{T_L}{J_m} \end{aligned} \quad (8)$$

To make the error dynamics of (8) strictly passive via feedback passivation, we choose $T_e$ as the input and $e_1$ as the output, and define the associated storage function and its time-derivative (symbolically) as follows

$$V_1 = V_1(e_1) \quad (9)$$

$$\dot{V}_1 = \dot{e}_1 \frac{\partial V_1}{\partial e_1} \quad (10)$$

Substituting (8) into (10), one obtains

$$\dot{V}_1 = \frac{\partial V_1}{\partial e_1} \left( \dot{\omega}_{rm}^* - \frac{T_e}{J_m} + \frac{\omega_{rm} B_m}{J_m} + \frac{T_L}{J_m} \right) \quad (11)$$

To make (8) an output strictly passive system, its input $T_e$ and $\partial V_1 / \partial e_1$ are chosen as

$$T_e = J_m \dot{\omega}_{rm}^* + T_L + B_m \omega_{rm} - v_1 + k_1 e_1 \quad (12)$$

$$\frac{\partial V_1}{\partial e_1} = J_m e_1 \quad (13)$$

where $v_1$ is a new input, and $k_1$ is a positive constant gain. Substituting (12) and (13) into (11), one gets

$$\dot{V}_1 = e_1 v_1 - k_1 e_1^2 \quad (14)$$

Integrating both sides of (14), we get the energy balance equation as

$$\int_0^t e_1 v_1 d\tau = V_1(t) - V_1(0) + \int_0^t k_1 e_1^2 d\tau \quad (15)$$

which implies that the error dynamics of speed tracking is output strictly passive with the new input $v_1$ [3]. Substitution of (12) into (8) with $v_1 = 0$, yields

$$J_m \dot{e}_1 + k_1 e_1 = 0 \quad (16)$$

### B. *Adaptive Feedback Passivation*

In practice, the values of the parameters in (12) cannot be measured precisely. In addition, the external load torque is unknown. As such, the parameter uncertainties and/or the unknown external disturbances would naturally degrade the system's overall performance. To deal with this problem, an adaptive feedback passivation approach is used. Let us rewrite the speed tracking error dynamics using (5) as

$$\begin{aligned} \dot{e}_1 &= \dot{\omega}_{rm}^* - \dot{\omega}_{rm} \\ &= \frac{1}{J_{mo}} \left( -T_e + J_{mo} \dot{\omega}_{rm}^* + (\omega_{rm}^* - e_1) B_m + (\hat{T}_L - \tilde{T}_L) - (\hat{F}_3 - \tilde{F}_3) \right) \\ &= -\frac{B_m}{J_{mo}} e_1 + \frac{1}{J_{mo}} \begin{pmatrix} -T_e + J_{mo} \dot{\omega}_{rm}^* + \omega_{rm}^* (\hat{B}_m - \tilde{B}_m) \\ + (\hat{T}_L - \tilde{T}_L) - (\hat{F}_3 - \tilde{F}_3) \end{pmatrix} \end{aligned} \quad (17)$$

where $\tilde{B}_m = \hat{B}_m - B_m$, $\tilde{T}_L = \hat{T}_L - T_L$, $\tilde{F}_3 = \hat{F}_3 - F_3$ and a variable with a hat denotes its estimated value, which is a notation adopted throughout the whole paper. Define the storage function $V_2$ and its time-derivative (symbolically) as follows

$$V_2 = V_2(e_1, \tilde{B}_m, \tilde{T}_L, \tilde{F}_3) \quad (18)$$

$$\dot{V}_2 = \dot{e}_1 \frac{\partial V_2}{\partial e_1} + \dot{\hat{B}}_m \frac{\partial V_2}{\partial \tilde{B}_m} + \dot{\hat{T}}_L \frac{\partial V_2}{\partial \tilde{T}_L} + \dot{\hat{F}}_3 \frac{\partial V_2}{\partial \tilde{F}_3} \quad (19)$$

Substituting (17) into (19), one obtains

$$\dot{V}_2 = \frac{\partial V_2}{\partial e_1}\frac{1}{J_{mo}}\left(-T_e + J_{mo}\dot{\omega}_{rm}^* + \omega_{rm}^*\hat{B}_m + \hat{T}_L - \hat{F}_3\right) - \frac{\partial V_2}{\partial e_1}\frac{B_m}{J_{mo}}e_1 + \left(\dot{\hat{B}}_m\frac{\partial V_2}{\partial \tilde{B}_m} - \frac{\omega_{rm}^*\tilde{B}_m}{J_{mo}}\frac{\partial V_2}{\partial e_1}\right) + \left(\dot{\hat{T}}_L\frac{\partial V_2}{\partial \tilde{T}_L} - \frac{\tilde{T}_L}{J_{mo}}\frac{\partial V_2}{\partial e_1}\right) + \left(\dot{\hat{F}}_3\frac{\partial V_2}{\partial \tilde{F}_3} + \frac{\tilde{F}_3}{J_{mo}}\frac{\partial V_2}{\partial e_1}\right) \quad (20)$$

To make (17) output strictly passive, we may specify $T_e$, $\partial V_2/\partial e_1$, $\partial V_2/\partial\tilde{B}_m$, $\partial V_2/\partial\tilde{T}_L$, and $\partial V_2/\partial\tilde{F}_3$ as follows:

$$T_e = J_{mo}\dot{\omega}_{rm}^* + \hat{T}_L + \omega_{rm}^*\hat{B}_m - \hat{F}_3 - v_2 + k_1 e_1 \quad (21)$$

$$\frac{\partial V_2}{\partial e_1} = J_{mo}e_1,\ \frac{\partial V_2}{\partial\tilde{B}_m} = \frac{\tilde{B}_m}{\gamma_1},\ \frac{\partial V_2}{\partial\tilde{T}_L} = \frac{\tilde{T}_L}{\gamma_2},\ \frac{\partial V_2}{\partial\tilde{F}_3} = \frac{\tilde{F}_3}{\gamma_3} \quad (22)$$

where $v_2$ is a new input, and $\gamma_1$, $\gamma_2$, and $\gamma_3$ are adaptation gains. From (20)-(22), we can get the following adaptation laws

$$\dot{\hat{B}}_m = \gamma_1\omega_{rm}^* e_1 \quad (23)$$

$$\dot{\hat{T}}_L = \gamma_2 e_1 \quad (24)$$

$$\dot{\hat{F}}_3 = -\gamma_3 e_1 \quad (25)$$

Substituting (21)-(25) into (20), one obtains

$$\dot{V}_2 = e_1 v_2 - \left(B_m + k_1\right)e_1^2 \quad (26)$$

Integrating both sides of (26), we get the energy balance equation as

$$\int_0^t e_1 v_2 d\tau = V_2(t) - V_2(0) + \int_0^t \left(B_m + k_1\right)e_1^2 d\tau \quad (27)$$

The new input $v_2$ can be designed to be of nonlinear type such that the system of (17) is asymptotically stable.

### C. Controller Design

The objectives are to combine the adaptive sliding-mode control with passivity property to maintain the speed on the desired trajectory. Unlike the input-output feedback linearization technique used in [8], the method in this paper does not require any nonlinear coordinate transformation. Besides, the proposed method is more robust than conventional adaptive backstepping controls [9] due to the fact that the sliding-mode control can effectively compensate for fast varying parameters in the system. In addition, the proposed controller can compensate for the uncertainties caused by the parameter variations and external load disturbance. Assume that the estimation errors $\tilde{B}_m$, $\tilde{T}_L$, and $\tilde{F}_3$ are bounded as $\left|\tilde{F}_3 - \tilde{T}_L\right| < \eta_1$ and $\left|\tilde{B}_m\right| < \eta_2$ where $\eta_1$ and $\eta_2$ are known. A sliding-mode control $v_2$ can be designed as

$$v_2 = -\eta_1\,\mathrm{sgn}(e_1) - \eta_2\left|\omega_{rm}^*\right|\mathrm{sgn}(e_1) \quad (28)$$

Using (21)-(25) and putting (28) into (20), one obtains

$$\dot{V}_2 = -\left(B_m + k_1\right)e_1^2 - \left(\eta_1 + \eta_2\left|\omega_{rm}^*\right|\right)\left|e_1\right| \le 0 \quad (29)$$

which results in a stable system. Increasing the gains $\eta_1$ and $\eta_2$ in (28), one can obtain more robustness. However, overly large values of the gains do lead to more chattering in the controls. To reduce the latter, a PI-saturation function is utilized to approximate the signum function [7]

$$\rho^*(e_1) = \begin{cases} \mathrm{sgn}(e_1), & if\ |e_1| > \Phi_1 \\ \dfrac{e_1}{\Phi_1} + \Gamma_1^*, & if\ |e_1| \le \Phi_1 \end{cases}, \quad (30)$$

where $\Phi_1$ is the thickness of the boundary layer, and $\Gamma_1^*$ is obtained by integration with respect to $e_1$. It is assumed that there exists such a value of $\Gamma_1^*$ with which the following inequalities hold when $e_1$ stays inside the boundary layer for a finite time [7]

$$\begin{cases} \dfrac{e_1}{\Phi_1} + \Gamma_1^* \ge 1, & for\ \dfrac{e_1}{\Phi_1} > 0 \\ \dfrac{e_1}{\Phi_1} + \Gamma_1^* \le -1, & for\ \dfrac{e_1}{\Phi_1} < 0 \end{cases} \quad (31)$$

In general, this value of $\Gamma_1^*$ is found by tuning. Given this, we may therefore use an adaptive law to approximate the $\Gamma_1^*$ value during a fixed sampling interval [5], [7]. Consider the following storage function and its time-derivative

$$V_3 = V_3\left(e_1, \tilde{B}_m, \tilde{T}_L, \tilde{F}_3, \tilde{\Gamma}_1\right)$$

$$\dot{V}_3 = \dot{e}_1\frac{\partial V_3}{\partial e_1} + \dot{\hat{B}}_m\frac{\partial V_3}{\partial\tilde{B}_m} + \dot{\hat{T}}_L\frac{\partial V_3}{\partial\tilde{T}_L} + \dot{\hat{F}}_3\frac{\partial V_3}{\partial\tilde{F}_3} + \dot{\hat{\Gamma}}_1\frac{\partial V_3}{\partial\tilde{\Gamma}_1} \quad (32)$$

where $\tilde{\Gamma}_1 = \hat{\Gamma}_1 - \Gamma_1^*$. According to previously results, we can choose $T_e$, $v_3$, $\dot{\hat{\Gamma}}_1$, $\partial V_3/\partial e_1$ and $\partial V_3/\partial\tilde{\Gamma}_1$ as follows:

$$T_e = J_{mo}\dot{\omega}_{rm}^* + \hat{T}_L + \omega_{rm}^*\hat{B}_m - \hat{F}_3 - v_3 + k_1 e_1 \quad (33)$$

$$v_3 = -\left(\eta_1 + \eta_2\left|\omega_{rm}^*\right|\right)\hat{\rho}(e_1) \quad (34)$$

$$\dot{\hat{\Gamma}}_1 = \gamma_4\left(\eta_1 + \eta_2\left|\omega_{rm}^*\right|\right)e_1 \quad (35)$$

$$\frac{\partial V_3}{\partial e_1} = J_{mo}e_1 \quad (36)$$

$$\frac{\partial V_3}{\partial\tilde{\Gamma}_1} = \frac{\tilde{\Gamma}_1}{\gamma_4} \quad (37)$$

where $\gamma_4$ is an adaptive gain, $v_3$ is a sliding-mode control, and $\hat{\rho}(e_1)$ is defined as

$$\hat{\rho}(e_1) = \begin{cases} \mathrm{sgn}(e_1), & if\ |e_1| > \Phi_1 \\ \dfrac{e_1}{\Phi_1} + \hat{\Gamma}_1, & if\ |e_1| \le \Phi_1 \end{cases} \quad (38)$$

Substituting (33) and (34) into (17), one can obtain

$$J_{mo}\dot{e}_1 + \left(k_1 + B_m\right)e_1 + \left(\eta_1 + \eta_2\left|\omega_{rm}^*\right|\right)\rho^*\left(e_1\right) = \tilde{u}_1 \quad (39)$$

where

$$\tilde{u}_1 = \Omega^T\tilde{\Theta} \quad (40)$$

$$\Omega^T = \left[-\omega_{rm}^* \quad -1 \quad 1 \quad -\left(\eta_1 + \eta_2\left|\omega_{rm}^*\right|\right)\right] \quad (41)$$

$$\tilde{\Theta}^T = \left[\tilde{B}_m \quad \tilde{T}_L \quad \tilde{F}_3 \quad \tilde{\Gamma}_1\right] \quad (42)$$

The closed-loop feedback interconnection for the speed control of the IPMSM is shown in Fig. 1. To prove the stability of the closed loop system, the passivity property of each block in Fig.1 is checked respectively. Using (39)-(42), we get

$$\langle -e_1, \tilde{u}_1 \rangle_t := \int_0^t -\tilde{u}_1^T e_1 d\tau$$
$$= \int_0^t \left( \tilde{T}_L - \tilde{F}_3 + \omega_{rm}^* \tilde{B}_m + \left( \eta_1 + \eta_2 \left| \omega_{rm}^* \right| \right) \tilde{\Gamma}_1 \right) e_1 d\tau$$
$$= \int_0^t \frac{1}{\gamma_1} \tilde{B}_m \dot{\tilde{B}}_m d\tau + \int_0^t \frac{1}{\gamma_2} \tilde{T}_L \dot{\tilde{T}}_L d\tau + \int_0^t \frac{1}{\gamma_3} \tilde{F}_3 \dot{\tilde{F}}_3 d\tau + \int_0^t \frac{1}{\gamma_4} \tilde{\Gamma}_1 \dot{\tilde{\Gamma}}_1 d\tau \quad (43)$$
$$= \frac{\tilde{B}_m^2(t) - \tilde{B}_m^2(0)}{2\gamma_1} + \frac{\tilde{T}_L^2(t) - \tilde{T}_L^2(0)}{2\gamma_2} + \frac{\tilde{F}_3^2(t) - \tilde{F}_3^2(0)}{2\gamma_3} + \frac{\tilde{\Gamma}_1^2(t) - \tilde{\Gamma}_1^2(0)}{2\gamma_4}$$

According to the definition [10], the upper block in Fig. 1 is passive from $-e_1$ to $\tilde{u}_1$. Next, we check the passivity of the lower block in Fig. 1. Utilizing (39), we obtain

$$\langle \tilde{u}_1, e_1 \rangle_t := \int_0^t e_1^T \tilde{u}_1 d\tau$$
$$= \int_0^t e_1^T \left( J_{mo} \dot{e}_1 + (k_1 + B_m) e_1 + \left( \eta_1 + \eta_2 \left| \omega_{rm}^* \right| \right) \rho^*(e_1) \right) d\tau \quad (44)$$
$$\geq \frac{J_{mo} e_1^2(t)}{2} - \frac{J_{mo} e_1^2(0)}{2} + \int_0^t (k_1 + B_m) e_1^2 d\tau$$

According to the definition [10], the lower block in Fig. 1 is strictly passive from $\tilde{u}_1$ to $e_1$. From (43) and (44), it is not difficult to obtain the following two equations

$$\left( \frac{\tilde{B}_m^2(t)}{2\gamma_1} + \frac{\tilde{F}_3^2(t)}{2\gamma_3} + \frac{\tilde{T}_L^2(t)}{2\gamma_2} + \frac{\tilde{\Gamma}_1^2(t)}{2\gamma_4} \right) = \left( \langle -e_1, \tilde{u}_1 \rangle_t + \frac{\tilde{B}_m^2(0)}{2\gamma_1} + \frac{\tilde{T}_L^2(0)}{2\gamma_2} + \frac{\tilde{F}_3^2(0)}{2\gamma_3} + \frac{\tilde{\Gamma}_1^2(0)}{2\gamma_4} \right) \quad (45)$$

$$\frac{J_{mo} e_1^2(t)}{2} \leq \langle \tilde{u}_1, e_1 \rangle_t + \frac{J_{mo} e_1^2(0)}{2} \quad (46)$$

Combining (45) and (46), we get

$$\frac{\tilde{B}_m^2(t)}{2\gamma_1} + \frac{\tilde{F}_3^2(t)}{2\gamma_3} + \frac{\tilde{T}_L^2(t)}{2\gamma_2} + \frac{\tilde{\Gamma}_1^2(t)}{2\gamma_4} + \frac{J_{mo} e_1^2(t)}{2} \leq \frac{\tilde{B}_m^2(0)}{2\gamma_1} + \frac{\tilde{T}_L^2(0)}{2\gamma_2} + \frac{\tilde{F}_3^2(0)}{2\gamma_3} + \frac{\tilde{\Gamma}_1^2(0)}{2\gamma_4} + \frac{J_{mo} e_1^2(0)}{2} \quad (47)$$

which implies $V_3(t) \leq V_3(0)$. Also, we can get [10]

$$\dot{V}_3 \leq -(k_1 + B_m) e_1^2 \leq 0 \quad (48)$$

Since $\dot{V}_3$ is negative semi-definite, it implies that $e_1$, $\tilde{F}_3$, $\tilde{\Gamma}_1$, $\tilde{B}_m$, and $\tilde{T}_L$ are bounded. By integrating (48), we can get

$$\int_0^\infty (k_1 + B_m) e_1^2 d\tau \leq -\int_0^\infty \dot{V}_3(\tau) d\tau = V_3(0) - V_3(\infty) < \infty \quad (49)$$

From (49), we can guarantee that $e_1 \in L_2$. As projection type of techniques can be used to ensure boundedness of estimated parameters during their estimation process, which in turn can ensure the boundedness of $T_e$; as a result, $\dot{e}_1$ is also bounded. Using previous results and Barbalat's lemma, we get the following

$$\lim_{t \to \infty} e_1(t) = 0 \quad (50)$$

From (50), we can conclude that the closed-loop system is asymptotically stable, given the fact that parameter uncertainties, external load disturbances, and estimation errors exist. Note that (50) does not imply that the estimated errors $\tilde{F}_3$, $\tilde{\Gamma}_1$, $\tilde{B}_m$, and $\tilde{T}_L$ will approach zero without satisfying the persistent excitation (PE) condition. With (33) being combined with (4), as currents generate torques in a motor, the passivity-based adaptive sliding-mode control law can be decomposed into d- and q-axis current commands as

$$i_d^* = \alpha \left| \Psi \right| \quad (51)$$

$$i_q^* = \frac{\Psi}{(3P_o/2)\lambda_m + (3P_o/2)(L_d - L_q)\alpha\left|\Psi\right|} \quad (52)$$

where

$$\Psi = T_e = J_{mo} \dot{\omega}_{rm}^* + \hat{T}_L + \hat{B}_m \omega_{rm}^* - \hat{F}_3 + \left( \eta_1 + \eta_2 \left| \omega_{rm}^* \right| \right) \hat{\rho}(e_1) + k_1 e_1 \quad (53)$$

where $\alpha$ is an adjustable parameter which must be negative and is used to make d- and q-axis currents satisfy the MTPA condition of the IPMSM. Recall that to achieve the MTPA, the d- and q-axis currents must satisfy the following equation

$$i_d^* = -\frac{\lambda_m}{2(L_d - L_q)} - \sqrt{\frac{\lambda_m^2}{4(L_d - L_q)^2} + i_q^{*2}} \quad (54)$$

Substituting (52) into (54) and using (51), we can get the d-axis current command as

$$i_d^* = -\frac{\lambda_m}{2(L_d - L_q)} - \sqrt{\frac{\lambda_m^2}{4(L_d - L_q)^2} + \left( \frac{i_d^*/\alpha}{(3P_o/2)\lambda_m + (3P_o/2)(L_d - L_q) i_d^*} \right)^2} \quad (55)$$

From (55), the parameter $\alpha$ can be obtained as

$$\alpha = \frac{i_d^*}{\sqrt{i_d^{*2} + \frac{\lambda_m i_d^*}{(L_d - L_q)}} \times \left( (3P_o/2)\lambda_m + (3P_o/2)(L_d - L_q) i_d^* \right)} \quad (56)$$

Using iterative method, $\alpha$ can be obtained in real-time implementation [11]. The whole system is shown in Fig. 2, which consists of a passivity-based adaptive sliding-mode controller, a parameter generator for $\alpha$, an IPMSM, and the parameter/external load estimators.

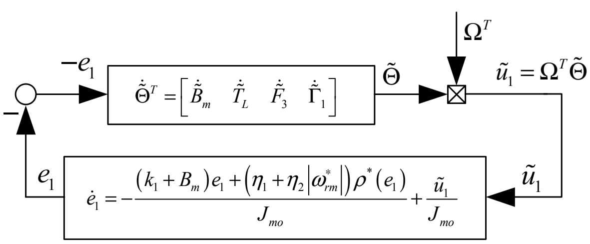



Fig. 1. Closed-loop feedback interconnection of the IPMSM.

## IV. Simulation Results

The effectiveness of the proposed passivity-based adaptive sliding-mode MTPA control is verified by numerical simulations. The parameters of the IPMSM used in this paper are shown in Table I. Simulations are conducted using three different test conditions. Comparison is made between two adaptive sliding-mode controllers. Both controllers are based on (21), (28), and (33)-(34).

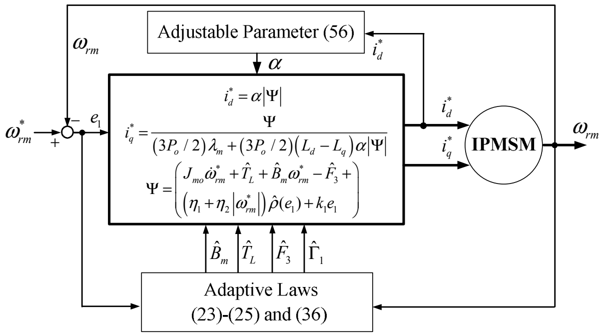
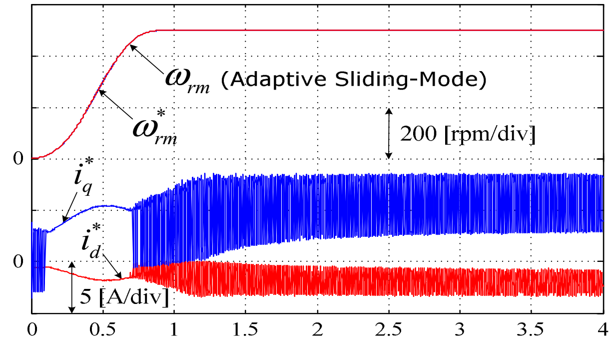



Fig. 2. Computation of the control input currents.

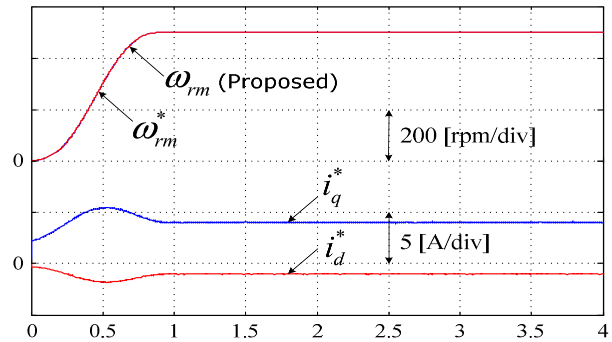
TABLE I
THE PARAMETERS OF IPMSM

| | |
|---|---|
| $P_o$ | 2 |
| $r_{so}$, Ω | 1.9 |
| $L_{do}$, mH | 15.1 |
| $L_{qo}$, mH | 31 |
| $\lambda_{mo}$, Vs/rad | 0.31 |
| $J_{mo}$ (without load), Kg-m$^2$ | 0.0005 |
| $B_{mo}$ (without load), Nms/rad | 0.03 |
| $J_{mo}$ (with load), Kg-m$^2$ | 0.0227 |
| $B_{mo}$ (with load), Nms/rad | 0.0341 |

The figures associated with the one without the PI-saturation function are designated as (a) whereas the one using the PI-saturation function, as proposed in this paper, as (b). The gains for the two control schemes are chosen to be the same as: $k_1 = 35$ , $\eta_1 = 1$ , $\eta_2 = 0.05$ , $\Phi_1 = 3$ , $\gamma_1 = 0.16$ , $\gamma_2 = 0.09$ , $\gamma_3 = 3.4$ , and $\gamma_4 = 15$ . Fig. 3(a) and 3(b) show the simulation results when there is a constant external load of 2 N.m. As one can observe, both controllers have good speed tracking performance. However, compared to Fig. 3(a), there is no chattering appears in Fig. 3(b). Fig. 4(a) and 4(b) show the simulation results of the two controllers tracking a periodic triangular-wave command without an external load. The undesirable chattering appears in Fig. 4(a) but not in Fig. 4(b). Obviously the proposed controller effectively attenuates the chattering occurs in the currents. Fig. 5(a) and 5(b) show the simulation results of the two controllers tracking a sinusoidal command with a load of 3.5 N.m. Comparing Fig. 5(a) and 5(b), one can see that the chattering from the adaptive sliding-mode control is largely removed when adaptive PI-saturation function is used. Although the adaptive sliding-mode controller can compensate lumped uncertainties and load disturbances, its switching function still caused unwanted chattering. Chattering seen in Fig. 5(b) is caused by the nature of inverter employing hysteresis current regulation strategy. From the simulation results we see that the proposed controller has significantly improved the performance of the adaptive sliding-mode controller for IPMSM drive systems.

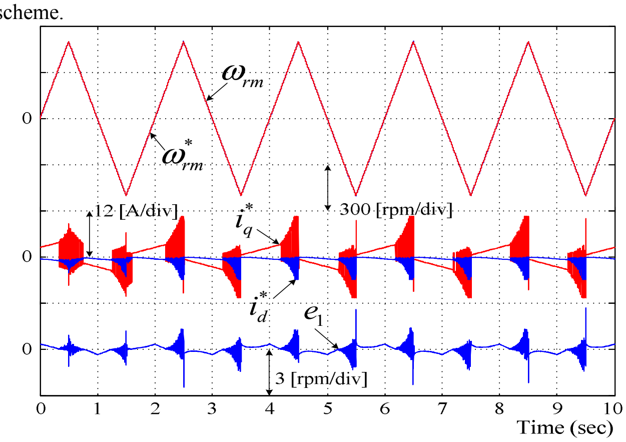



(a)

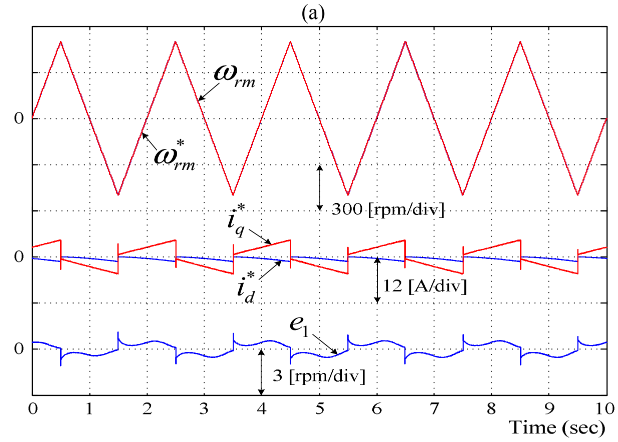



(b)

Fig. 3. Speed response and d-q axis current responses at 500 rpm with 2 N.m load (a) adaptive sliding-mode control scheme (b) proposed control scheme.

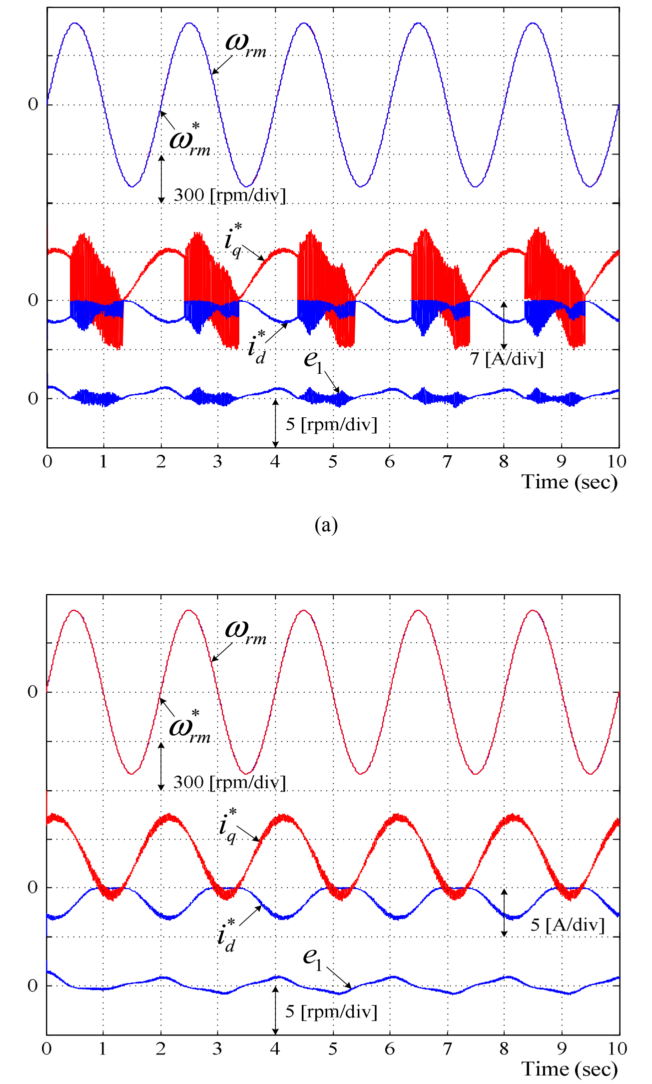



(a)




(b)

Fig. 4. Speed response, d-q axis current responses, and speed tracking error response of periodic triangular-wave command (a) adaptive sliding-mode control scheme (b) proposed control scheme.




(a)

(b)

Fig. 5. Speed response, d-q axis current responses, and speed tracking error response of periodic sinusoidal-wave command with 3.5 N.m load (a) adaptive sliding-mode control scheme (b) proposed control scheme.

## V. CONCLUSIONS

In this paper, a passivity-based adaptive sliding-mode controller for speed tracking is proposed for IPMSM drive systems. First, the feedback passivation approach is used to make the tracking error dynamics a strictly passive system with a designated input. After that, we applied an adaptive sliding-mode technique to obtain the current control input to the motor, which is further decomposed into d- and q-axis current commands that satisfy the MTPA condition. To reduce chattering, we employed an adaptive PI saturation function to approximate the signum function within boundary layer. Simulation results showed satisfactory performance and verified numerically the effectiveness of the proposed scheme.


## ACKNOWLEDGMENT

This research was supported by the National Science Council of the Republic of China under grant NSC 97-2221-E-002-178-MY3.


## REFERENCES


[1] L. U. Gökdere and M. A. Simaan, "A passivity-based method for induction motor control," *IEEE Trans. Ind. Electron.*, vol. 44, no. 5, pp. 688-695, Oct. 1997.

[2] W. J. Wang and J. Y. Chen, "Compositive adaptive position control of induction motors based on passivity theory," *IEEE Trans. Energy Conver.*, vol. 16, no. 2, pp. 180-185, June 2001.

[3] W. J. Wang and J. Y. Chen, "Passivity-based sliding mode position control for induction motor drives," *IEEE Trans. Energy Conver.*, vol. 20, no. 2, pp. 316-321, June 2005.

[4] C. I. Byrnes, A. Isidori, and J. C. Willems, "Passivity, feedback equivalence, and the global stabilization of minimum phase nonlinear systems," *IEEE Trans. Automat. Contr.*, vol. 36, no. 11, pp. 1228-1240, Nov. 1991.

[5] F. J. Lin, C. K. Chang, and P. K. Huang, "FPGA-based adaptive backstepping sliding-mode control for linear induction motor drive," *IEEE Trans. Power Electron.*, vol. 22, no. 4, pp. 1222-1231, July 2007.

[6] Y. Xu, "Chattering free robust control for nonlinear systems," *IEEE Trans. Control Syst. Technol.*, vol. 16, no. 6, pp. 1352-1359, Nov. 2008.

[7] R. Shahnazi, H. M. Shanechi, and N. Pariz, "Position control of induction and DC servomotors: a novel adaptive fuzzy PI sliding mode control" *IEEE Trans. Energy Conver.*, vol. 23, no. 1, pp. 138-147, Mar. 2008.

[8] C. K. Lin, T. H. Liu, and S. H. Yang, "Nonlinear position controller design with input-output linearisation technique for an interior permanent magnet synchronous motor control system," *IET Power Electr.*, vol. 1, no. 1, pp. 14-26, Mar. 2008.

[9] M. A. Rahman, D. M. Vilathgamuwa, N. Uddin, and K. J. Tseng, "Nonlinear control of interior permanent-magnet synchronous motor," *IEEE Trans. Ind. Applicat.*, vol. 39, no. 2, pp. 408-416, Mar./Apr. 2003.

[10] M. Krstic, I. Kanellakopoulos, and P. Kokotovic, "Nonlinear and adaptive control design," New York: John Wiley & Sons, Inc., 1995.

[11] J. L. Shi, T. H. Liu, and Y. C. Chang, "Adaptive controller design for a sensorless IPMSM drive system with a maximum torque control," *IEE Proc.- Electr. Power Appl.*, vol. 153, no. 6, pp. 823-833, Nov. 2006.