Nonlinear Control of Interior Permanent-Magnet Synchronous Motor

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***Abstract*—This paper presents a novel speed control technique for an interior permanent-magnet synchronous motor (IPMSM) drive based on newly developed adaptive backstepping technique. The proposed stabilizing feedback law for the IPMSM drive is shown to be globally asymptotically stable in the context of Lyapunov theory. The adaptive backstepping technique takes system nonlinearities into account in the control system design stage. The detailed derivations of the control laws have been given for controller design. The complete IPMSM drive incorporating the proposed backstepping control technique has been successfully implemented in real-time using digital signal processor board DS1102 for a laboratory 1-hp motor. The performance of the proposed drive is investigated both in experiment and simulation at different operating conditions. It is found that the proposed control technique provides a good speed tracking performance for the IPMSM drive ensuring the global stability.**

***Index Terms*—Adaptive backstepping, interior permanent-magnet synchronous motor (IPMSM), nonlinear control, real-time implementation.**

I. Introduction

THE advances in power semiconductor technology, digital electronics, magnetic materials, and control theory have enabled modern ac motor drives to face challenging high-efficiency and high-performance requirements in the industrial sector. Among ac drives, the permanent-magnet synchronous motor has been gaining popularity owing to its high torque to current ratio, large power to weight ratio, high efficiency, high power factor, and robustness. These features are due to the incorporation of high-energy rare-earth alloys such as neodymium–iron–boron in its construction. In particular, the interior permanent-magnet synchronous motor (IPMSM) which has magnets buried in the rotor core exhibits certain good properties, such as mechanically robust rotor construction, rotor physical nonsaliency of the air gap, and small effective air gap. The rotors of these permanent magnet motors have complex geometry to ensure optimal use of the expensive permanent magnet material while maintaining a high magnetic field in the air gap. These features allow the IPMSM drive to be operated in high-speed mode by incorporating the field-weakening technique.

Usually, high-performance motor drives require fast and accurate response, quick recovery from any disturbances, and insensitivity to parameter variations. The dynamic behavior of an ac motor can be significantly improved using vector control theory where motor variables are transformed into an orthogonal set of d–q axes such that speed and torque can be controlled separately [1]. This gives the IPMSM machine the highly desirable dynamic performance capabilities of the separately excited dc machine, while retaining the general advantages of the ac over dc motors. Originally, vector control was applied to the induction motor and a vast amount of research work has been devoted to this area. The vector control method is relevant to the IPMSM drive as the control is completely carried out through the stator, as the rotor excitation control is not possible.

However, precise speed control of an IPMSM drive becomes a complex issue owing to nonlinear coupling among its winding currents and the rotor speed as well as the nonlinearity present in the torque equation. The system nonlinearity becomes severe if the IPMSM drive operates in the field weakening region where the direct axis current $i_d \neq 0$. This results in the appearance of a nonlinear term, which would have vanished under the existing vector control scheme with $i_d = 0$.

There have been significant developments in nonlinear control theory applicable to electric motor drives. Interestingly, the d–q transformation applicable to ac motors can be considered as a feedback linearization transformation. However, with the recent developments in nonlinear control theories, a modern control engineer has not only found a systematic approach in dealing with nonlinearities but has managed to develop approaches which had not been considered previously. The surge of such nonlinear control methods applicable to electromechanical systems include variable-structure systems [2], differential geometric approach [3], [4], and passivity theory [5].

As electrical drive systems possess well-defined nonlinear model characteristics, they have become good candidates for the application of newly developed nonlinear control techniques. If the knowledge of such nonlinearities can be included in the design of nonlinear controller, an enhanced dynamic behavior of the motor drive can be accomplished. One ingenious way of

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designing a nonlinear controller is adaptive backstepping (AB) [6]–[8]. According to this control technique, tracking objectives designed using nonlinear control theory in the mechanical subsystem of the IPMSM drive. Therefore, the reference currents of the IPMSM motor is developed to fulfill necessary torque (or speed) requirements. Subsequent control voltages of the IPMSM drive is then derived to force actual current to follow reference values thus effectively meeting the motion control objective which is embedded inside current tracking objectives. The tracking controllers described in this paper fall into the category of a full state-feedback controller with system uncertainty. In this scheme, the system uncertainties and unknown bounded disturbances are taken into account. The adaptive backstepping control technique based on full state-feedback is designed to overcome system model uncertainties. This is especially important to the IPMSM drive as electrical parameters of the IPMSM drive tend to vary with the load current [9] as well as with the operating conditions such as temperature and saturation [9], [10]. To the best of the authors' knowledge, there has not been any systematic stability analysis of the IPMSM drive reported in the literature. In this paper we illustrate the effectiveness of adaptive backstepping technique applicable to the IPMSM drive. Using Lyapunov stability-type techniques one may guarantee the global stability of the IPMSM drive and nonuniform speed tracking performance.

This paper is organized as follows. Firstly, the IPMSM drive model in the d–q reference frame is presented in Section II. In Section III, adaptive-backstepping control methodology is developed. This method effectively compensates for the IPMSM system uncertainties. Section IV describes the experimental implementation of the proposed controller for a laboratory 1-hp IPMSM using a digital signal oricessor (DSP) board DS1102. Finally, sample results are presented in both simulation and experiment at different operating conditions in order to verify the efficacy and stability of the proposed controller.

II. IPMSM Drive Model

The mathematical model of a IPMSM drive can be described by the following equations in a synchronously rotating rotor d–q reference frame as [11]:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R + pL_d & -P\omega_r L_q \\ P\omega_r L_d & R + pL_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ P\omega_r \psi_f \end{bmatrix} \tag{1}$$

$$T_e = T_L + J_m p\omega_r + B_m \omega_r \tag{2}$$

$$T_e = \frac{3P}{2}[\psi_f i_q + (L_d - L_q) i_d i_q] \tag{3}$$

where

v_d, v_q	d- and q-axes stator voltages;
i_d, i_q	d- and q-axes stator currents;
R	stator per-phase resistance;
L_d, L_q	d- and q-axes stator inductances;
T_e, T_L	electromagnetic and load torques;
J_m	moment of inertia of the motor and load;
B_m	friction coefficient of the motor;
P	number of poles of the motor;
ω_r	rotor speed in angular frequency;
p	differential operator ($= d/dt$);
ψ_f	rotor magnetic flux linking the stator.

It is well known that a synchronous motor is unable to self-start when supplied with a constant frequency source. The starting torque in the IPMSM drive used in this research is provided by a rotor squirrel cage winding. The starting process of the IPMSM drive can be considered as a superposition of two operating modes, namely, unsymmetric asynchronous motor mode and magnet-excited asynchronous generator mode [9]. Therefore, the effect of shorted rotor windings has to be considered, if one wants to examine the process of run-up to the synchronization. However, the model equations in (1)-(3) do not describe the asynchronous behavior of the IPMSM drive.

The IPMSM drives constructed using neodymium–iron–boron magnets can be operated over a wide temperature range [10]. It has been shown that, within normal operating temperature range, the residual flux density and intrinsic coercivity will decrease as the temperature is increased. However, this is considered as a reversible process as the temperature comes down to normal value, the flux density and coercivity will approximately return to their original values. This variation in residual flux along with the stator resistance, in turn, affects the dynamic behavior of the IPM motor.

The standard linear d–q axes IPMSM model with constant parameters will lead to an unsatisfactory prediction of the performance of an interior permanent-magnet motor owing to the extraordinary saturation of these machines during normal operation. It has been shown that improved prediction of IPMSM behavior can be accomplished by adjusting the model parameters according to the changing saturating conditions [12]. Various researchers have proved that there exist variations in X_d, X_q and ψ_f with the direct and quadratic axis saturation as well as with the direction of rotation [9], [12]. In light of these findings we propose to use adaptive backstepping technique wherein system parameter variations will be taken into account at the design stage of the controller.

III. Control Design

The objective of this paper is to obtain the IPMSM control voltages in order to achieve high-performance speed tracking. According to the motor model given in (1)–(3) of Section II, it can be seen that the speed control can be achieved by controlling the q-axis component v_q of the supply voltage as long as the d-axis current i_d is maintained at zero. This results in the electromagnetic torque being directly proportional to the current i_q. Since $i_d = 0$, the d-axis flux linkage depends only on the rotor permanent magnets. The resultant IPMSM model can be represented as

$$pi_q = \frac{1}{L_q}(v_q - Ri_q - P\omega_r \psi_f) \tag{4}$$

$$v_d = -\omega_r L_q i_q \tag{5}$$

$$T_e = T_L + J_m p\omega_r + B_m \omega_r \tag{6}$$

$$T_e = \frac{3P}{2}(\psi_f i_q). \tag{7}$$

A. Backstepping Design for IPMSM

The proposed control strategy adopted in this paper is called adaptive backstepping. This recently developed nonlinear control technique has been gaining popularity among control system engineers [13]. This control technique can be effectively used to linearize a nonlinear system in the presence of uncertainties. Unlike in other feedback linearization methods, the advantage of adaptive backstepping is its flexibility whereby useful nonlinearities can be kept intact during stabilization [6]. The essence of adaptive backstepping is the identification of a virtual control state and force it to become a stabilizing function. Thus, it generates a corresponding error variable [6]. For second order systems, this error variable can be stabilized by proper selection of control input via Lyapunov stability theory.

For the IPMSM drive, the pertinent control objective is to track the rotor angular speed. The tracking error can be given as

$$e = \omega_r^* - \omega_r \tag{8}$$

and speed error dynamic is given by

$$J\dot{e} = J\dot{\omega}_r = B_m\omega_r + T_L - \frac{3P}{2}\left[\psi_f i_q + (L_d - L_q) i_d i_q\right]. \tag{9}$$

As the speed error needs to be reduced to zero, the d–q-axes current components i_d and i_q are identified as the virtual control elements to stabilize the motor speed.

To determine the stabilizing function, the following Lyapunov function is defined as [14]:

$$V = \frac{1}{2}e^2. \tag{10}$$

By differentiating the Lyapunov function (10), one easily obtains

$$\begin{aligned}\dot{V} &= e\dot{e} = (B_m\omega_r + T_L)\frac{e}{j} - \frac{3P}{2J}\left[\dot{\psi}_f i_q + (L_d - L_q) i_d i_q\right] e \\ &= -k_s e^2 + \frac{e}{j}\left(B_m\omega_r + T_L - \frac{3P}{2}\psi_f i_q + k_s J e\right) \\ &\quad - \frac{3P}{2J}(L_d - L_q) i_d i_q e \end{aligned} \tag{11}$$

where k_s is the closed-loop feedback constant. The speed control tracking is achieved if one defines the following stabilizing functions:

$$i_q^* = \frac{2}{3P\psi_f}(B_m\omega_r + T_L + k_s J e) \tag{12}$$

$$i_d^* = 0. \tag{13}$$

If one makes the current tracking errors $e_d = i_d^* - i_d$ and $e_q = i_q^* - i_q$ to be zero by virtue of selecting proper control voltages, the Lyapunov function (10) becomes

$$\dot{V} = -k_s e^2 \tag{14}$$

thus, achieving global asymptotical stability.

As load torque T_L is unknown, it has to be estimated adaptively. Thus, let us define

$$\hat{i}_q^* = \frac{2}{3P\psi_f}(B_m\omega_r + \hat{T}_L + k_s J e) \tag{15}$$

where $\hat{T}_L$ is the estimated value of the load torque. Thus, from (9) and (15), the following speed error dynamics are obtained as:

$$\dot{e} = \frac{1}{J}\left[-\tilde{T}_L + \frac{3P}{2}\psi_f e_q - k_s e J + \frac{3P}{2}(L_d - L_q) e_d i_q\right] \tag{16}$$

where $\tilde{T}_L = \hat{T}_L - T_L$.

To stabilize the current components i_d and i_q, one defines the following d–q axes current error dynamics as:

$$\dot{e}_d = \dot{i}_d^* - \dot{i}_d = \frac{R}{L_d} i_d - \frac{P\omega_r L_q}{L_d} i_q - \frac{v_d}{L_d} \tag{17}$$

$$\begin{aligned}\dot{e}_q &= \dot{i}_q^* - \dot{i}_q = \frac{2}{3P\psi_f}\left(B_m\frac{d\omega_r}{dt} + k_s J\frac{de}{dt}\right) - \frac{di_q}{dt} \\ &= \frac{2}{3P\psi_f}(B_m - k_s J)\frac{d\omega_r}{dt} \\ &\quad - \left(-\frac{Ri_q}{L_q} - \frac{P\omega_r L_d}{L_q} i_d + \frac{v_q}{L_q} - \frac{P\omega_r\psi_f}{L_q}\right) \\ &= \frac{2(B_m - k_s J)}{3P\psi_f J}\left[\frac{3P}{2}(\psi_f i_q + (L_d - L_q) i_d i_q) - B\omega_r - T_L\right] \\ &\quad + \frac{Ri_q}{L_q} + \frac{P\omega_r L_d}{L_q} i_d - \frac{v_q}{L_q} + \frac{P\omega_r\psi_f}{L_q}. \end{aligned} \tag{18}$$

As we know the motor inductances L_d and L_q vary with the saturation of the stator and rotor cores, thus, it is necessary to estimate them adaptively. Therefore, let these estimates be $\hat{L}_d$ and $\hat{L}_q$. Now, one can define a new Lyapunov function including the d–q axes current error variables e_d and e_q and estimation error variables as

$$V_2 = \frac{1}{2}\left(e^2 + e_d^2 + e_q^2 + \frac{1}{\gamma_1}\tilde{L}_d^2 + \frac{1}{\gamma_2}\tilde{L}_d^2 + \frac{1}{\gamma_1}\tilde{T}_L^2\right) \tag{19}$$

where $\tilde{L}_d = \hat{L}_d - L_d$, $\tilde{L}_q = \hat{L}_q - L_q$, and γ_1, γ_2, and γ_3 are adaptive gains.

By differentiating the Lyapunov function V_2 in (19) one easily obtains (20), as shown at the bottom of the next page. If the following d–q axes control voltages are selected:

$$\begin{aligned} v_d &= Ri_d - P\omega_r\hat{L}_q i_q + k_1 e_d\hat{L}_d \\ &\quad + \frac{3P}{2J}\hat{L}_d(\hat{L}_d - \hat{L}_q) i_q e \end{aligned} \tag{21}$$

$$\begin{aligned} v_q &= \frac{2\hat{L}_q(B_m - k_s J)}{3P\psi_f J}\left(\frac{3P}{2}\left[\psi_f i_q + (\hat{L}_d - \hat{L}_q) i_d i_q\right]\right. \\ &\quad \left. - B\omega_r - \hat{T}_L\right) \\ &\quad + Ri_q + P\omega_r\hat{L}_d i_d + P\omega_r\psi_f + k_2 e_q\hat{L}_q \\ &\quad + \frac{3P}{2J}\psi_f e\hat{L}_q. \end{aligned} \tag{22}$$

Further simplifying (20), one obtains

$$\begin{aligned} \dot{V}_2 &= \frac{e}{J}\left[\widetilde{T}_L + \frac{3P}{2}\psi_f e_q - k_s e J + \frac{3P}{2}(L_d - Lq) e_d i_q\right] \\ &\quad + \frac{e_d}{\hat{L}_d}\left[-P\omega_r 8(L_q - \hat{L}_q) i_q - k_1 e_d\hat{L}_d \right. \\ &\qquad \left. - \frac{3P}{2J}(\hat{L}_d - \hat{L}_q)\hat{L}_d e i_q\right] \\ &\quad + \frac{e_n}{\hat{L}_q}\begin{bmatrix} \frac{\hat{L}_q(B_m - k_s J)}{\psi_f J}(L_d - \hat{L}_d - L_q + \hat{L}_d) i_d i_q \\ -\frac{2\hat{L}_q(B_m - k_s J)}{3P\psi_f J}(T_L - \hat{T}_L) + P\omega_r(L_d - \hat{L}_d) i_d \\ -k_2 e_q\hat{L}_q - \frac{3P}{2J}\psi_f e\hat{L}_q \end{bmatrix} \\ &\quad + \frac{1}{\gamma_1}\widetilde{L}_d\dot{\widetilde{L}}_d + \frac{1}{\gamma_2}\widetilde{L}_q\dot{\widetilde{L}}_q + \frac{1}{\gamma_3}\widetilde{T}_L\dot{\widetilde{T}}_L. \end{aligned} \tag{23}$$

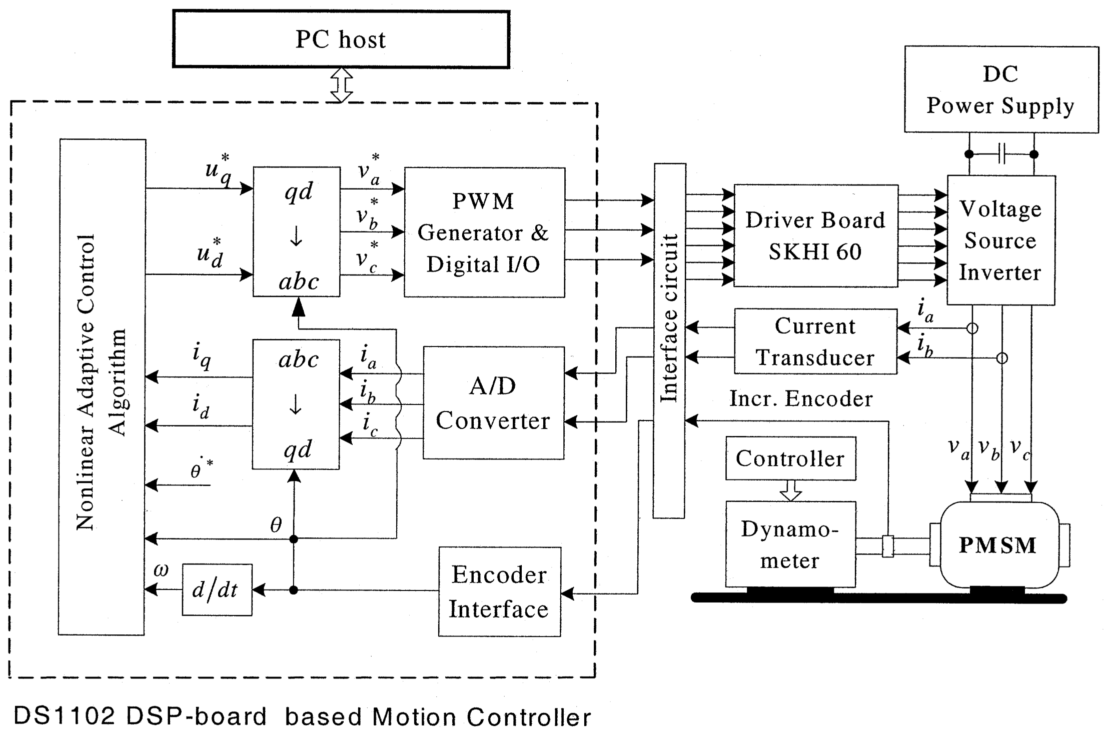


Fig. 1. Control block diagram of the IPMSM drive.

Equation (23) results in the following expression:

$$\begin{aligned}\dot{V}_2 =& -k_s e^2 - k_1 e_d^2 - k_2 e_q^e \\ &+ \tilde{T}_L \left[\frac{e}{j} + \frac{2e_q(B_m - k_s J)}{3P\psi_f J} + \frac{1}{\gamma_3}\dot{\tilde{T}}_L\right] \\ &+ \tilde{L}_d \left[-\frac{3Pee_d i_q}{2J} - \frac{e_q(B_m - k_s J)i_d e_q}{\psi_f J} \right. \\ &\qquad \left. -\frac{e - qP\omega_r i_d}{L_q} + \frac{1}{\gamma_1}\dot{\tilde{L}}_d\right] \\ &+ \tilde{L}_q \left[\frac{3Pee_d i_q}{2J} + \frac{e_d P\omega_r i_q}{L_d} \right. \\ &\qquad \left. + \frac{e_q(B_m - k_s J)i_d i_q}{\psi_f J} + \frac{1}{\gamma_2}\dot{\tilde{L}}_1\right]. \end{aligned} \tag{24}$$

From (24), the following update laws can be derived as:

$$\dot{\hat{T}}_L = -\gamma_3 \left[\frac{e}{J} + \frac{2e_q(B_m - k_s J)}{3P\psi_f J}\right] \tag{25}$$

$$\begin{aligned}\dot{\hat{L}}_d =& -\gamma_1 \left[-\frac{3Pee_d i_q}{2J} - \frac{e_q(B_m - k_s J)i_d i_q}{\psi_f J}\right. \\ &\qquad \left. -\frac{e_q P\omega_r i_d}{L_q}\right]\end{aligned} \tag{26}$$

$$\begin{aligned}\dot{\hat{L}}_q =& -\gamma_2 \left[\frac{3Pee_d i_q}{2J}\right. \\ &\qquad \left. +\frac{e_d P\omega_r i_q}{L_d} + \frac{e_q(B_m - k_s J)i_d i_q}{\psi_f J}\right].\end{aligned} \tag{27}$$

$$\begin{aligned}\dot{V}_2 =& e\dot{e} + e_d\dot{e}_d + e_q\dot{e}_q + \frac{1}{\gamma_1}\tilde{L}_d\dot{\tilde{L}}_d + \frac{1}{\gamma_2}\tilde{L}_q\dot{\tilde{L}}_q + \frac{1}{\gamma_3}\tilde{T}_L\dot{\tilde{T}}_L = -k_s e^2 - k_1 e_d^2 - k_2 e_q^2 \\ &+ \frac{e}{J}\left[-\tilde{T}_L + \frac{3P}{2}\psi_f e_q - k_s eJ + \frac{3P}{2}(L_d - L_q)e_d i_q\right] \\ &+ e_d\left(\frac{R}{L_d}i_d - \frac{P\omega_r L_q}{L_d}i_q - \frac{v_d}{L_d} + k_1 e_d\right) \\ &+ e_q\left\{\begin{array}{l}\dfrac{2(B_m - k_s J)}{3P\psi_f J}\left(\dfrac{3P}{2}\left[\psi_f i_q + (L_d - L_q)i_d i_q\right] - B\omega_r - T_L\right) \\ \qquad + \dfrac{Ri_q}{L_q} + \dfrac{P\omega_r L_d}{L_q}i_d - \dfrac{v_q}{L_q} + \dfrac{P\omega_r\psi_f}{L_q} + k_2 e_q\end{array}\right\} \\ &+ \frac{1}{\gamma_1}\tilde{L}_d\dot{\hat{L}}_d + \frac{1}{\gamma_2}\tilde{L}_q\dot{\hat{L}}_q + \frac{1}{\gamma_3}\tilde{T}_L\dot{\hat{T}}_L.\end{aligned} \tag{20}$$

TABLE I
MACHINE PARAMETERS

Motor rated power	3-phase, 1 *hp*
Rated voltage	208 *V*
Rated current	3 *A*
Rated frequency	60 *Hz*
Pole pair number (*P*)	2
d-axis inductance, L_d	42.44 *mH*
q-axis inductance, L_q	79.57 *mH*
Stator resistance, *R*	1.93 Ω
Motor inertia, J_m	0.003 kgm^2
Friction coefficient, B_m	0.001 *Nm/rad/s*
Magnetic flux constant, ψ_f	0.311 *volts/rad/s*

Therefore, one can obtain the following expression for the derivative of the Lyapunov function as:

$$\dot{V}_2 = -k_s e^2 - k_1 e_d^2 - k_2 e_q^2 \leq 0. \tag{28}$$

It is proved that (28) guarantees the global asymptotic stability in the current loop. Real-time implementation of the drive control system based on the above control algorithm is addressed next.

IV. REAL-TIME IMPLEMENTATION

The proposed backstepping nonlinear control technique for IPMSM is experimentally implemented using DSP board DS1102 through both hardware and software [15]. The DSP board is installed in a PC with uninterrupted communication capabilities through dual-port memory. The hardware schematic for real-time implementation of the proposed FLC-based IM drive is shown in Fig. 1. The DS1102 board is based on a Texas Instruments (TI) TMS320C31 32-bit floating-point DSP. The DSP has been supplemented by a set of on-board peripherals used in digital control systems, such as A/D, D/A converters and incremental encoder interfaces. The DS1102 is also equipped with a TI TMS320P14 16-bit microcontroller DSP that acts as a slave processor and is used for some special purposes. In this work, a slave processor is used for digital I/O configuration. The actual motor currents are measured by the Hall-effect sensors which have good frequency response and fed to the DSP board through A/D converter. As the motor neutral is isolated, only two phases currents are fed back and the other phase current is calculated from them. The rotor position is measured by an optical incremental encoder which is mounted at the rotor shaft end. Then, it is fed to the DSP board through encoder interface. The encoder generates 4096 pulses per revolution. By using a fourfold pulse multiplication the number of pulses is increased to 4×4096 in order to get better resolution. A 24-bit position counter is used to count the encoder pulses and is read by a calling function in the software.

The motor speed is calculated from the rotor position by backward difference interpolation. A digital moving average filter is used to remove the noise from the speed signal. The calculated actual motor speed is used to calculate the torque component of

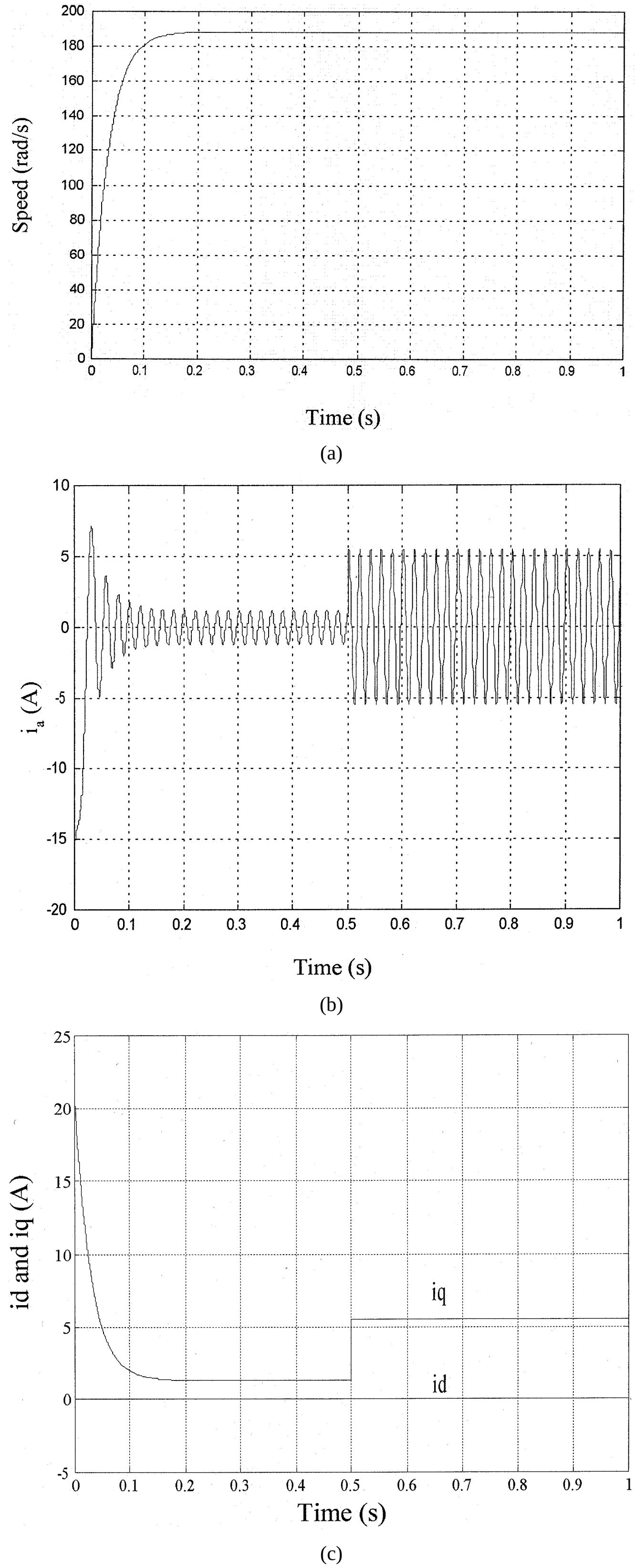


Fig. 2. Simulated drive responses for a sudden increase in load for the proposed control system. (a) Speed. (b) Stator current i_a. (c) i_d, i_q.

the current i_q^* according to (12). The load torque T_L is updated according to (25). Then, v_d and v_q are calculated using (21) and (22), respectively. The three phase voltages are obtained from

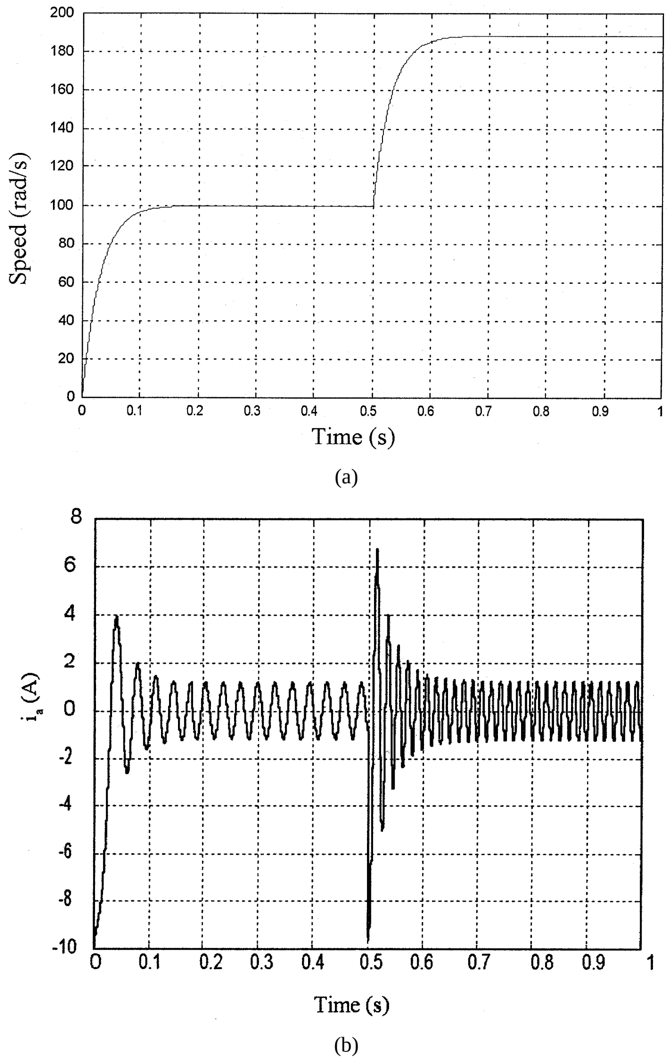


Fig. 3. Simulated drive responses for a sudden increase in speed for the proposed control system. (a) Speed. (b) Stator current i_a.

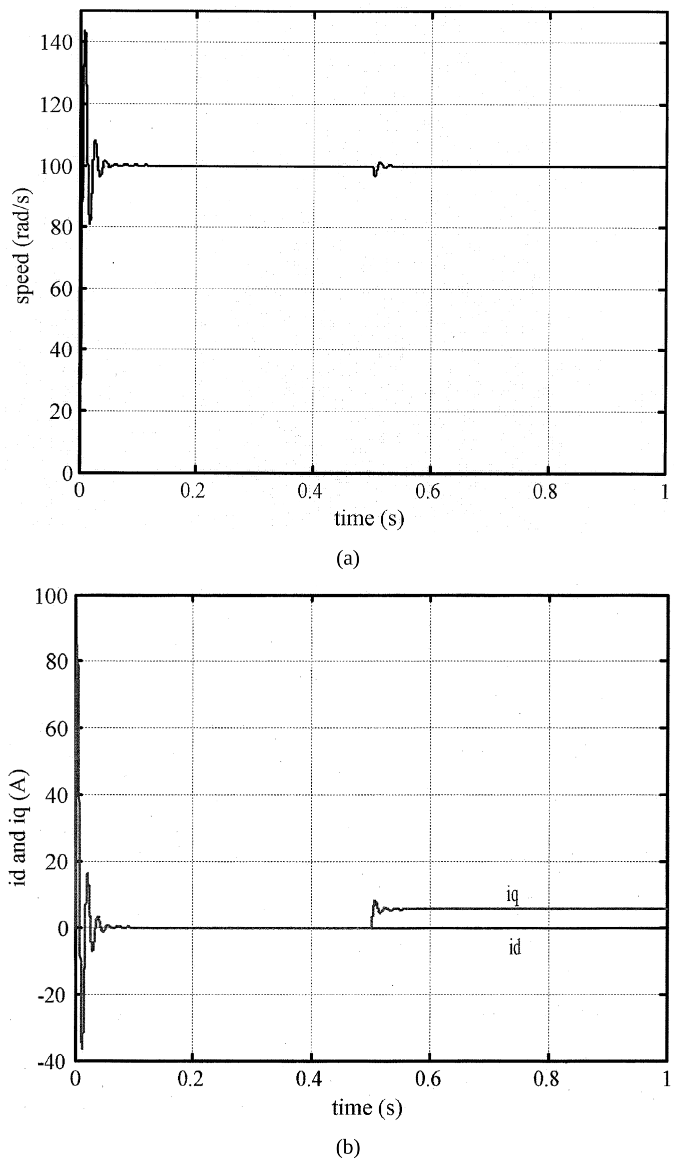


Fig. 4. Simulated drive responses for a sudden increase in load for a PI-controlled system. (a) Speed. (b)Stator currents i_d, i_q.

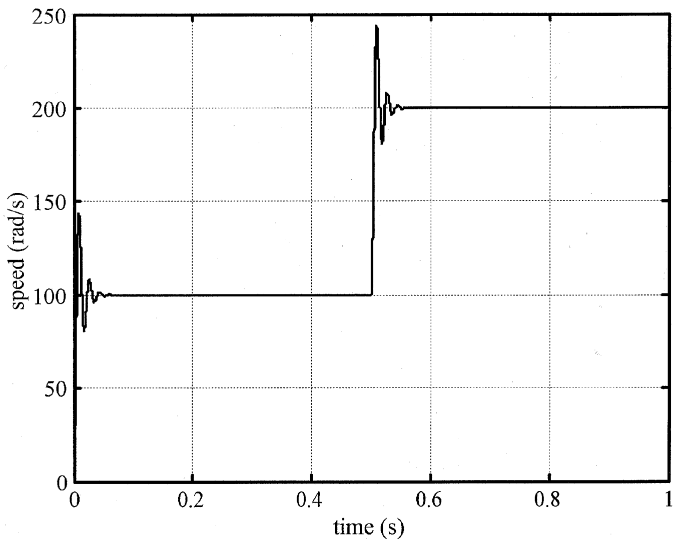
d and q axes components using inverse Park's transformation [11]. These phase voltages are compared with a high-frequency carrier signal to generate the pulsewidth-modulated (PWM) signals, which act as firing pulses for the six transistors of the inverter. Thus, these six PWM logic signals are the output of the DSP board and fed to the base drive circuit of the inverter power module. The D/A channels are used to capture the necessary output signals in a digital storage oscilloscope.

The complete IPMSM drive is implemented through software by developing a program in high-level ANSI "C" programming language. The program is compiled by the TI "C" compiler and then the program is downloaded to the DSP controller board. The sampling frequency for experimental implementation of the proposed IPMSM drive system is 10 kHz.

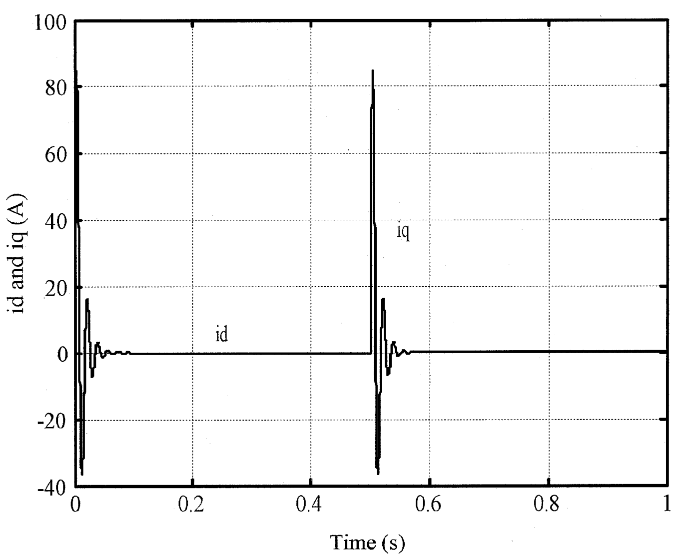
V. Results and Discussion

In order to validate the mathematical analysis and, hence, to establish the effectiveness of the proposed backstepping control scheme, the performance of the IPMSM drive based on the proposed control scheme is investigated both theoretically and experimentally at different operating conditions. Sample results are presented below.

Digital simulations have been carried out using MATLAB/SIMULINK [16]. The parameters of the laboratory IPMSM drive are given in Table I. The simulated motor speed and current responses for the proposed controller are shown in Fig. 2(a)–(c) to see the starting performance as well as the sudden load impact. The drive system is started at a constant load of 1N·m with the speed reference set at 1800 r/min (188.5 rad/s). It can be seen from Fig. 2(a) that the actual speed converges to the reference value within 0.2 s. However, according to Fig. 2(b) and (c), the stator current shows an overshoot but it lasts for only 0.15 s. At $t = 0.5\ s$, a load torque of 5 N·m is applied to the motor shaft in a stepwise manner. The actual speed does not change during the disturbance while the stator current swiftly reaches to its new value corresponding to the load applied. This shows the capability of new controller in terms of disturbance rejection. Another simulated speed

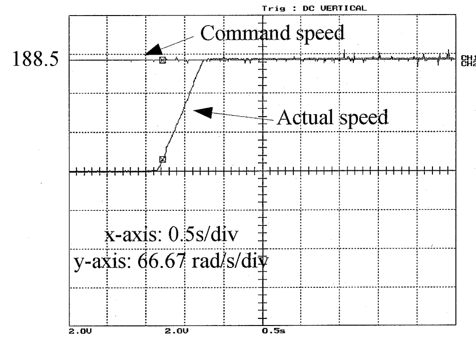


(a)

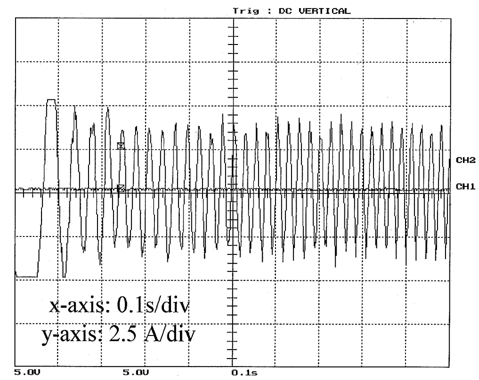


(b)

Fig. 5. Simulated drive responses for a sudden increase in speed for a PI-controlled system. (a) Speed. (b) Stator currents i_d, i_q.



(a)



(b)

Fig. 6. Experimental starting performance of the proposed drive. (a) Speed. (b) Stator current i_a.

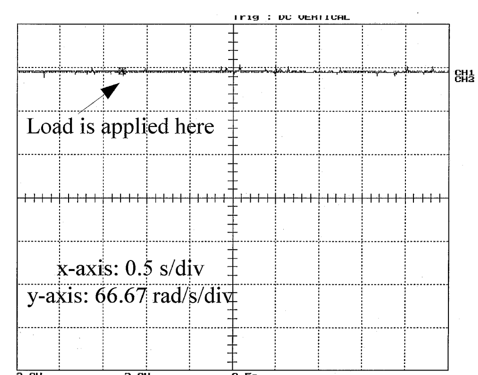


Fig. 7. Experimental speed response of the proposed IPMSM drive for a sudden increase in load.

response for a sudden increase in command speed is shown in Fig. 3. It is evident from this figure that the proposed backstepping control technique is also capable of handling the disturbance in speed command. Computer simulations have been carried out to determine system responses for an industry standard proportional-plus-integral (PI)-controlled IPMSM where PI controllers are employed in the speed as well as in the current loops. Control parameters of this control system are given in the Appendix. Fig. 4 shows the speed and current transients during starting and for a load disturbance at $t = 0.5s$. Both speed and q-axis current exhibit large overshoots during starting before the speed settles to its steady-state value of 100 rad/s. These overshoots are significantly larger than those in the adaptive backstepping controller. The settling time is around 1 s. At the beginning of the load disturbance, the q-axis current component shows a significant overshoot from its steady-state value while speed shows a transient drop of 2–3 rad/s. Fig. 5 shows the current and speed responses for a sudden increase in the reference speed. The q-axis current as well as the speed shows remarkable transients during transition. Although steady-state error of the speed is quite low, the overall performance is inferior to that of the proposed controller.

The experimental starting performance including speed and stator current i_a are shown in Fig. 6(a) and (b), respectively. It is shown that the proposed drive is also capable of following the command speed very quickly with zero steady-state error and without any overshoot or undershoot in a real-time situation. Fig. 7 shows the experimental speed response due to a sudden increase in load. It is quite evident that there is virtually no change of speed due to load variation. Another experimental speed response is shown in Fig. 8 for a sudden change in command speed. It is evident from Fig. 8 that the proposed drive can adapt itself with speed disturbance.

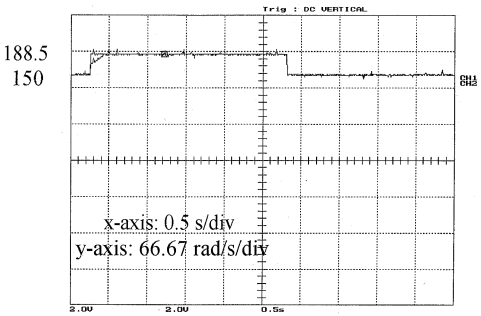


Fig. 8. Experimental speed response of the proposed IPMSM drive for a sudden change in command speed.

VI. Conclusions

Successful application of adaptive backstepping control for the speed tracking of the IPMSM drive has been illustrated in this paper. It has been shown that the IPMSM drive belongs to a class of nonlinear system for which adaptive backstepping technique can be effectively used. By recursive manner, virtual control states of the IPMSM drive have been identified and stabilizing control laws are developed subsequently using Lyapunov stability theory. The detailed derivation for the control laws has been provided. The complete IPMSM drive has been successfully implemented in real-time using DSP board DS1102 for the laboratory 1-hp motor. The validity of the proposed control technique has been established both in simulation and experiment at different operating conditions.

Appendix

The following parameters are used in the PI controllers employed in the speed and current control loops. Speed control loop: $k_p = 70$, $k_i = 0.1$. Current control loops: $k_p = 0.5$, $k_i = 0.001$.

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