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Review Article

ADAPTIVE SPEED CONTROL FOR SURFACE MOUNTED PERMANENT MAGNET SYNCHRONOUS MOTOR DRIVE USING RENEWABLE ENERGY SYSTEM

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ABSTRACT

This paper proposes an adaptive proportional integral derivative (PID) speed control scheme for permanent magnet synchronous motor (PMSM) drives. The proposed controller consists of three control terms: a decoupling term, a PID term, and a supervisory term. The first control term is employed to compensate for the nonlinear factors, the second term is made to automatically adjust the control gains, and the third one is designed to guarantee the system stability. Different from the offline-tuning PID controllers, the proposed adaptive controller includes adaptive tuning laws to online adjust the control gains based on the gradient descent method. Thus, it can adaptively deal with any system parameter uncertainties in reality. The proposed scheme is not only simple and easy to implement, but also it guarantees an accurate and fast speed tracking. It is proven that the control system is asymptotically stable. To confirm the effectiveness of the proposed algorithm, the comparative experiments between the proposed adaptive PID controller and the conventional PID controller are performed on the PMSM drive. Finally, it is validated that the proposed design scheme accomplishes the superior control performance (faster transient response and smaller steady-state error) compared to the conventional PID method in the presence of parameter uncertainties.

INTRODUCTION

IN recent years, the ac motors are extensively applied in home appliances as well as industrial applications such as electric vehicles, wind generation systems, industrial robots, air conditioners, washing machines, etc. There are two main categories of the ac motors: induction motors (IMs) and permanent magnet synchronous motors (PMSMs). Nowadays, the IMs are used in about 70% of industrial electric motors due to their simplicity, ruggedness, and low production costs. Despite that, the PMSMs are gradually taking over the IMs owing to their high efficiency, low maintenance cost, and high power density. However, the PMSM system is not easy to control because it is a nonlinear multivariable system and its performance can be highly affected by parameters variations in the run time. Therefore, researchers always desire to design a high-performance controller which has a simple algorithm, fast response, high accuracy, and robustness against the motor parameter and load torque variations. Traditionally, the proportional–integral–derivative (PID) controller is widely adopted to control the PMSM systems in industrial applications owing to its simplicity, clear functionality, and effectiveness. However, a big problem of the traditional PID controller is its sensitivity to the system uncertainties. Thus, the control performance of the conventional PID method can be seriously degraded under parameter variations. Some groups of researchers try to overcome this disadvantage by proposing the hybrid PID controllers or new tuning rules. A hybrid control system, which contains a fuzzy controller in the transient and a PI controller in the steady-state, is proposed. In the fuzzy rules are employed for tuning the PI gains. Unfortunately, both these methods use offline-tuning rules, which lack the adaptability to deal with the time-varying system uncertainties. An adaptive PI controller with an online-tuning rule is presented in . Although this controller does not require the exact knowledge of any motor parameter, the authors do not show the results under parameter uncertainties. Recently, many researchers have presented various advanced control strategies to efficiently control the PMSM systems, such as fuzzy logic control (FLC), nonlinear optimal control (NOC), sliding mode control (SMC), neutral network control (NNC), adaptive control, etc. The FLC is a preferred research topic due to its fuzzy reasoning capacity.

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However, as the number of the fuzzy rules increases, the control accuracy can get better but the control algorithm can be complex. The NOC is successfully applied on the PMSM drives. Unfortunately, this control method requires full knowledge of the motor parameters with a sufficient accuracy and the results under serious variations of the mechanical parameters are not shown. The SMC has achieved much popularity in the speed control of the PMSM drives because of its great properties such as robustness to external load disturbances and fast dynamic response. However, its system dynamics are still subject to the parameter variations and chattering problem. Meanwhile, the NNC technique has been presented as a substitutive design method to control the speed of the PMSM system. Next, the adaptive control is also an interesting method for the PMSM drives because it can deal with the motor parameter and load torque variations.

Nevertheless, in these two papers, only the stator inductances and load torque variations are considered. The authors neglect the uncertainties of other motor parameters such as stator resistance, moment of inertia and viscous friction coefficient, etc. Moreover, the adaptive control algorithm does not guarantee the convergence condition of the system dynamic error. By combining the simplicity and effectiveness of the traditional PID control and the automatic adjustment capability of the adaptive control, this paper proposes a simple adaptive PID control algorithm for the PMSM drives. The adaptive PID controller encompasses the adaptive tuning laws which are designed to online adjust the control gains by using the supervisory gradient descent method. Therefore, when the motor parameters vary, the PID gains are automatically tuned to attain their optimal values. Consequently, the proposed control system achieves a good regulation performance such as fast dynamic response and small steady-state error even under system parameter uncertainties. The stability analysis of the proposed control strategy is described in details through the Lyapunov stability theories. The experimental results demonstrate the validity and feasibility of the proposed adaptive PID control method in comparison with the conventional PID control scheme under parameter uncertainties.

System Model Description and Dynamic Error System

System Model Description

The mathematical model of a surface-mounted permanent magnet synchronous motor (SPMSM) drives can be described by the following equations in a *d*-*q*

$$\begin{aligned}\dot{\omega} &= \frac{3}{2J}\frac{p^2}{4}\psi_m i_{qs} - \frac{B}{J}\omega - \frac{p}{2J}T_L \\ \dot{i}_{qs} &= -\frac{R_s}{L_s}i_{qs} - \frac{\psi_m}{L_s}\omega + \frac{1}{L_s}V_{qs} - \omega i_{ds} \\ \dot{i}_{ds} &= -\frac{R_s}{L_s}i_{ds} + \frac{1}{L_s}V_{ds} + \omega i_{qs}\end{aligned} \tag{1}$$

synchronously rotating reference frame:

Where *ω* is the electrical rotor speed, *ids* and *iqs* are the *d*-axis and *q*-axis stator currents, *Vds* and *Vqs* are the *d*-axis and *q*-axis voltage inputs, *Ls* is the stator inductance, *Rs* is the stator resistance, *ψm* is the magnetic flux linkage, *J* is the moment of inertia, *B* is the viscous friction coefficient, *p* is the number of poles, and *TL* is the load torque. Depending on *Rs*, *Ls*, *J*, *B*, and *ψm*, the system parameters *k*1~*k*6 can be denoted asThen the SPMSM drive system model is rewritten as the following

$$\begin{aligned}k_1 &= \frac{3}{2J}\frac{p^2}{4}\psi_m,\ k_2 = \frac{B}{J},\ k_3 = \frac{p}{2J} \\ k_4 &= \frac{R_s}{L_s},\ k_5 = \frac{\psi_m}{L_s},\ k_6 = \frac{1}{L_s}\end{aligned} \tag{2}$$

equations:

Then the SPMSM drive system model is

$$\begin{aligned}\dot{\omega} &= k_1 i_{qs} - k_2\omega - k_3 T_L \\ \dot{i}_{qs} &= -k_4 i_{qs} - k_5\omega + k_6 V_{qs} - \omega i_{ds} \\ \dot{i}_{ds} &= -k_4 i_{ds} + k_6 V_{ds} + \omega i_{qs}\end{aligned} \tag{3}$$

rewritten as the following equations:

Conventional PID Controller with Decoupling Term First, the speed error (*ωe*) and rotor acceleration (*β*) are defined as

$$\begin{aligned}\omega_e &= \omega - \omega_d \\ \beta &= \dot{\omega} = k_1 i_{qs} - k_2\omega - k_3 T_L\end{aligned} \tag{4}$$

Where *ωd* is the desired speed. From (3) and (4), the following dynamic equations can be derived

$$\begin{aligned}\dot{\omega}_e &= \dot{\omega} - \dot{\omega}_d \\ \dot{\beta} &= k_1(-k_4 i_{qs} - k_5\omega + k_6 V_{qs} - \omega i_{ds}) - k_2\dot{\omega} - k_3\dot{T}_L \\ \dot{i}_{ds} &= -k_4 i_{ds} + k_6 V_{ds} + \omega i_{qs}\end{aligned} \tag{5}$$

In practical applications, the desired speed and the load torque vary slowly in the sampling period. Thus, it can be

Reasonably supposed that the derivatives of *ωd* and *TL* can be neglected. Then the system model (1) can be rewritten as

$$\begin{aligned}\dot{\omega}_e &= \beta \\ \dot{\beta} &= -k_2\beta - k_1k_4 i_{qs} - k_1k_5\omega - k_1\omega i_{ds} + k_1k_6 V_{qs} \\ \dot{i}_{ds} &= -k_4 i_{ds} + \omega i_{qs} + k_6 V_{ds}\end{aligned} \tag{6}$$

Then, the second-order system can be achieved in the following:

$$\begin{aligned}\ddot{\omega}_e + \lambda\dot{\omega}_e &= -k_2\beta - k_1k_4 i_{qs} - k_1k_5\omega - k_1\omega i_{ds} + \lambda\beta \\ \dot{i}_{ds} &= -k_4 i_{ds} + \omega i_{qs} + k_6 V_{ds}\end{aligned} \tag{7}$$

Where λ is the positive control parameter. Based on the basic theory of the feedback linearization control, the decoupling control term *uf* = [*u*1*f u*2*f*]*T* is chosen as

$$\begin{aligned}u_{1f} &= (k_1k_4 i_{qs} + k_1k_5\omega + k_1\omega i_{ds} + (k_2 - \lambda)\beta)/(k_1k_6), \\ u_{2f} &= (k_4 i_{ds} - \omega i_{qs})/k_6\end{aligned} \tag{8}$$

From (7) and (8), the dynamic error system can be formulated as follows:

$$\dot{\omega}_e = -\lambda\omega_e + k_1k_6(V_{qs} - u_{1f})$$
$$i_{ds} = k_6(V_{ds} - u_{2f}) \quad (9)$$

Then the conventional PID controller is given by

Where B = diag [$k1k6$, $k6$], uf is the decoupling control

$$V_{dqs} = \begin{bmatrix} k_1k_6V_{qs} \\ k_6V_{ds} \end{bmatrix} = B\begin{bmatrix} V_{qs} \\ V_{ds} \end{bmatrix} = u_f + u_{PID} \quad (10)$$

term to compensate for the nonlinear factors, and $uPID$ is the PID Control term below.

Where ($K1P$, $K2P$), ($K1I$, $K2I$), and ($K1D$) are the proportional gains, integral gains, and derivative gain of

$$u_{PID} = \begin{bmatrix} u_{1PID} \\ u_{2PID} \end{bmatrix} = \begin{bmatrix} -K_{1P}\,\omega_e - K_{1I}\int_0^t \omega_e dt - K_{1D}\dfrac{d\omega_e}{dt} \\ -K_{2P}\,i_{dse} - K_{2I}\int_0^t i_{dse}dt \end{bmatrix} = EK \quad (11)$$

the PID control term, respectively. The state and gain matrices are given as

$$E = \begin{bmatrix} \int_0^t \omega_e dt & \omega_e & \beta & 0 & 0 \\ 0 & 0 & 0 & \int_0^t i_{ds}dt & i_{ds} \end{bmatrix},$$
$$K = [-K_{1I} \quad -K_{1P} \quad -K_{1D} \quad -K_{2I} \quad -K_{2P}]^T$$

From top to bottom, the estimated position and the estimation error are given, respectively. At the beginning

Proposed Adaptive Pidcontroller Design

The conventional PID controller with the offline-tuned control gains can give a good control performance if the motor parameters ($k1$ to $k6$) are accurately known. However, the system parameters gradually change during operating time; therefore, after a long running time, the control performance can be seriously degraded if changed system parameters are not updated.

To overcome this challenge, this section presents the adaptive tuning laws for auto adjustment of the control gains. On that note, the control gains, denoted as $K1I$, $K1P$, $K1D$, $K2I$, and $K2P$ in are adjusted to the proper values based on the supervisory gradient descent method. The proposed adaptive PID controller is assumed to have the following form:

$$V_{dqs} = u_f + u_{PID} - u_{PID0} + u_S \quad (12)$$

Supervisory control term which guarantees the system stability and $uPID0$ $= EK_0$ (with $K_0 = [K_{1I0} \; K_{1P0} \; K_{1D0} \; K_{2I0} \; K_{2P0}]^T$ is a Constant coefficient matrix).

Proposed Adaptive PID Controller

In order to derive the proper adaptation laws, a new tracking error vector based on the reduced-order sliding mode dynamics is defined as

$$s(t) = \begin{bmatrix} s_1(t) \\ s_2(t) \end{bmatrix} = \begin{bmatrix} \lambda\omega_e + \beta \\ i_{ds} \end{bmatrix} \quad (13)$$

Then, the transfer function G (p) from $s1$ to ω_e is given by the following strictly positive real function:

$$G(p) = \frac{\omega_e}{s_1} = \frac{1}{(\lambda + p)} \quad (14)$$

Where p is the Laplace variable. Hence, it can be concluded that ω_e converges to zero as $s \to 0$. From the viewpoints of the SMC method, the sliding condition that ensures the hitting and existence of a sliding mode is deduced according to the Lyapunov stability theory. Commonly, the Lyapunov function candidate for the sliding mode control is given by $V1$ = $sTs/2$. Then, the sliding condition can be obtained from the Lyapunov stability theory as

$$\dot{V}_1(t) = s^T\dot{s} < 0 \quad (15)$$

The gradient descent search algorithm is calculated in the direction opposite to the energy Flow, and the convergence properties of the PID gains can also be obtained. Therefore, the adaptation laws for the five control gains $K1P$, $K1I$, $K1D$, $K2P$, and $K2I$ can be easily obtained based on the supervisory gradient method as

$$\begin{aligned}
\dot{K}_{1P} &= -\gamma_{1P}\frac{\partial V_1}{\partial K_{1P}} = -\gamma_{1P}\frac{\partial V_1}{\partial u_{1PID}}\frac{\partial u_{1PID}}{\partial K_{1P}} = -\gamma_{1P}s_1\,\omega_e \\
\dot{K}_{1I} &= -\gamma_{1I}\frac{\partial V_1}{\partial K_{1I}} = -\gamma_{1P}\frac{\partial V_1}{\partial u_{1PID}}\frac{\partial u_{1PID}}{\partial K_{1I}} = -\gamma_{1I}s_1\int_0^t \omega_e dt \\
\dot{K}_{1D} &= -\gamma_{1D}\frac{\partial V_1}{\partial K_{1D}} = -\gamma_{1D}\frac{\partial V_1}{\partial u_{1PID}}\frac{\partial u_{1PID}}{\partial K_{1D}} = -\gamma_{1D}\,s_1\beta \\
\dot{K}_{2P} &= -\gamma_{2P}\frac{\partial V_1}{\partial K_{2P}} = -\gamma_{2P}\frac{\partial V_1}{\partial u_{2PID}}\frac{\partial u_{2PID}}{\partial K_{2P}} = -\gamma_{2P}\,s_2\,i_{ds} \\
\dot{K}_{2I} &= -\gamma_{2I}\frac{\partial V_1}{\partial K_{2I}} = -\gamma_{2I}\frac{\partial V_1}{\partial u_{2PID}}\frac{\partial u_{2PID}}{\partial K_{2I}} = -\gamma_{2I}s_2\int_0^t i_{ds}dt
\end{aligned} \quad (16)$$

Where $\gamma 1P$, $\gamma 1I$, $\gamma 1D$, $\gamma 2P$, and $\gamma 2I$ are the positive learning-rates. The adaptive tuning laws can be expressed in the following vector form:

$$\dot{K} = \Phi E^T s \quad (17)$$

Where Φ = diag ($\gamma 1I$, $\gamma 1P$, $\gamma 1D$, $\gamma 2I$, $\gamma 2P$).

Remark 1: By utilizing the online-tuning rules the control gains are automatically adjusted as the system parameters vary. Therefore, the proposed adaptive PID controller can overcome the disadvantage of all offline-tuning methods and can exhibit the good performance regardless of the system parameter uncertainties. Next, the supervisory control term in necessary for pulling back the dynamic errors to the predetermined bounded region and guaranteeing the system stability. Assume that there exists an optimal PID control term (*u*PID*) such that

$$u^*_{PID} = u_{PID0} + \varepsilon \quad (18)$$

Then the supervisory control term is designed as

$$u_S = \begin{bmatrix} -\delta_1 \cdot \mathrm{sgn}(s_1) \\ -\delta_2 \cdot \mathrm{sgn}(s_2) \end{bmatrix} \quad (19)$$

To this end, the desired controller is obtained by combining the decoupling control term (8), PID control term with the adaptation laws (16), and supervisory control term as *Vdqs* = *uf* + *uPID* + *us*.

Stability Analysis To analyze the stability of the dynamic error system, the following theorem is established.

Theorem: Consider the dynamic error system represented if the adaptive PID speed controller with the adaptive tuning laws is applied to then, the dynamic error system is asymptotically stable.

Proof: The following equation can be derived from

$$\dot s = B(u_{PID} + u_S - u^*_{PID} + u^*_{PID} - u_{PID0}) = B(E\tilde K + u_S - \varepsilon) \quad (20)$$

follows:

where $\tilde K = K - K^*$

Let us define the errors of the control gains as follows:

$$\tilde K_{1P} = K_{1P} - K^*_{1P}, \tilde K_{1I} = K_{1I} - K^*_{1I}, \tilde K_{1D} = K_{1D} - K^*_{1D}, \tilde K_{2P} = K_{2P} - K^*_{2P}, \tilde K_{2I} = K_{2I} - K^*_{2I} \quad (21)$$

Based on (20) and (21) the following Lyapunov function candidate is chosen:

$$V_2(t) = \frac{1}{2}s^T B^{-1} s + \frac{1}{\gamma_{1P}}\tilde K^2_{1P} + \frac{1}{\gamma_{1I}}\tilde K^2_{1I} + \frac{1}{\gamma_{1D}}\tilde K^2_{1D} + \frac{1}{\gamma_{2P}}\tilde K^2_{2P} + \frac{1}{\gamma_{2I}}\tilde K^2_{2I} \quad (22)$$

The time derivative of the Lyapunov function *V*2 (*t*) is given by

$$\begin{aligned} \dot V_2(t) - s^T B^{-1}\dot s &+ \frac{1}{\gamma_{1I}}\tilde K_{1I}\dot K_{1I} + \frac{1}{\gamma_{1P}}\tilde K_{1P}\dot K_{1P} \\ &+ \frac{1}{\gamma_{1D}}\tilde K_{1D}\dot K_{1D} + \frac{1}{\gamma_{2I}}\tilde K_{2I}\dot K_{2I} + \frac{1}{\gamma_{2P}}\tilde K_{2P}\dot K_{2P} \\ &- s^T B^{-1} B(E\tilde K + u_S - \kappa) + \tilde K^T \Phi^{-1}\dot{\tilde K} \\ &= s^T\left(E\tilde K - \begin{bmatrix}\delta_1 \,\mathrm{sgn}(s_1) \\ \delta_2 \,\mathrm{sgn}(s_2)\end{bmatrix} - \varepsilon\right) - \tilde K^T \Phi^{-1}\Phi E^T s \\ &= -\delta_1|s_1| - \delta_2|s_2| - \varepsilon_1 s_1 - \varepsilon_2 s_2 \\ &\le -(\delta_1 - |\varepsilon_1|)|s_1| - (\delta_2 - |\varepsilon_2|)|s_2| \\ &\le 0 \end{aligned} \quad (23)$$

Using the adaptive PID controller (12) with the adaptive tuning laws (16), the inequality $\dot V_2(t) \le 0$ can be obtained for non-zero value of the tracking error vector *s*. Since $\dot V_2(t)$ is a negative semi-definite function (i.e., *V*2(*t*) ≤ *V*2(0)), which implies that *s* and *K*~ are bounded. Let the function $\Omega(t) = [(\delta_1 - |\gamma_1|)|s_1| + (\delta_2 - |\gamma_2|)|s_2|]$ and the following inequality is obtained from (23)

$$\int_0^t \Omega(\tau)\,d\tau \le V_2(0) - V_2(t) \quad (24)$$

Because *V*2 (0) is bounded and *V*2 (*t*) is bounded and non-increasing, thus the following inequality can be deduced Meanwhile, as far as *s* is bounded, by using (13), it is obvious to realize that *s* is also bounded.

$$\lim_{t\to\infty}\int_0^{\infty} \Omega(\tau)\,d\tau < \infty \quad (25)$$

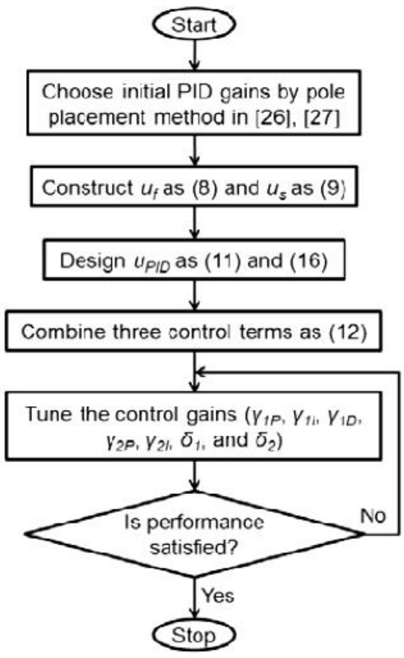
Meanwhile, as far as *s* is bounded, by using (13), it is obvious to realize that *s* is also bounded.

Remark 2: The angular acceleration (*β*) is normally not available. This angular acceleration can be estimated by the extended state observer.

However, these estimations require accurate knowledge about some system parameters, so the algorithm seems to be complex and the accuracy of estimated values is highly sensitive to parameters variations. In this paper, *β* is simply computed by using the relation $\beta = \omega$ as Moreover, this computation is independent of the system parameters.

$$\hat\beta(k) - \frac{\varphi}{T+\varphi}\hat\beta(k-1) + \frac{1}{T+\varphi}\left[\omega(k) - \omega(k-1)\right] \quad (26)$$

Where *φ* is a sufficiently small filter time constant to limit the vulnerability of this computation to noise.

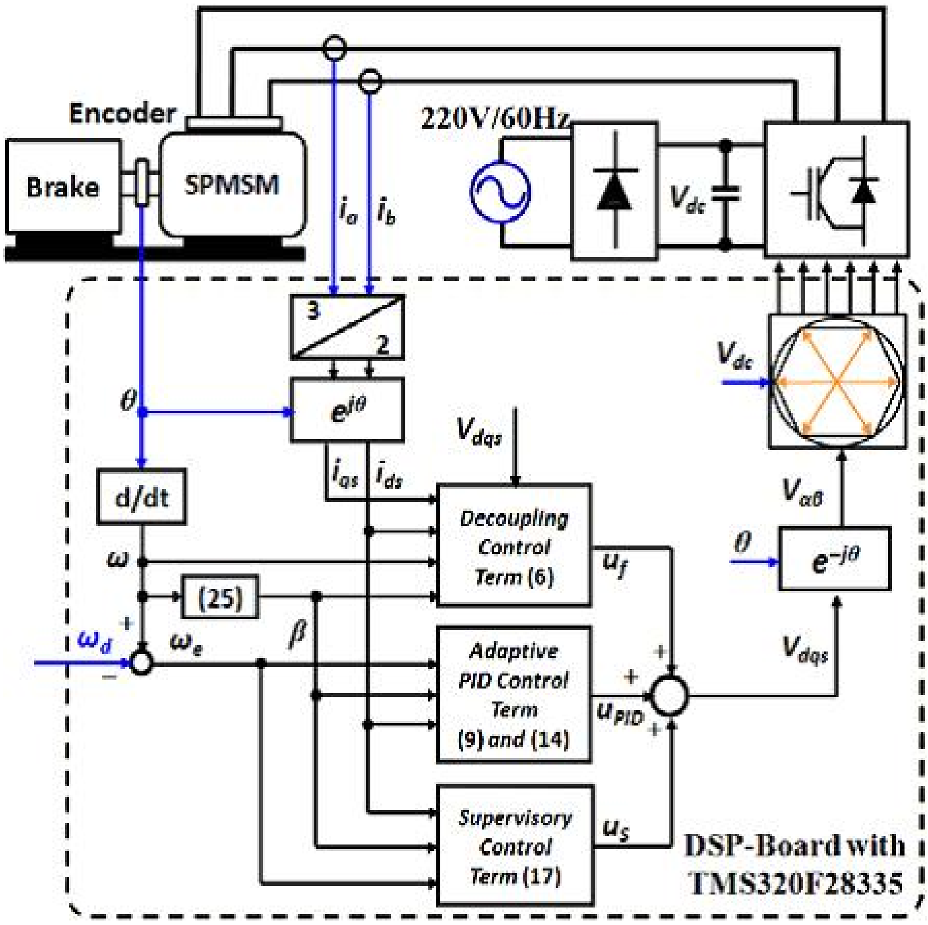


Remark 3: It should be noted that the proposed adaptive PID control strategy can be applicable to various electrical systems which have the mathematical form as (9). The overall design procedure of the proposed control scheme can be summarized as follows:

Step 1: Choosing the initial values of the PID gains by using the pole placement method.
Step 2: Constructing the decoupling control term *uf* as (8) and the supervisory control term *uS* as.
Step 3: Making the PID control term *uPID* as with the adaptation laws.
Step 4: Giving the desired controller by combining the three control terms.
The proposed adaptive PID control algorithm that represents the design procedure described in *Remark 3*.

Fig. 2 illustrates the overall block diagram of the SPMSM drive system with the proposed adaptive PID controller which consists of three control terms.

As the driving system includes the following hardware components: a three-phase inverter and its driving circuits, a SPMSM, a brake, an incremental encoder, Hall-effect current sensors, and a DSP-board. First, the rotor position is measured via the encoder RIA-40–2500ZO and then it is used to calculate the motor speed. Also, the transformations such as Clarke, Park, and inverse Park are utilized to transform a stationary three-phase system into the stationary two-phase system (*abc*-frame to *αβ*-frame), the stationary system to the synchronously rotating system (*αβ*-frame to *dq*-frame), and the synchronously rotating system to the stationary system (*dq*-frame to *αβ*-frame), respectively.



Experimental Evaluation

Drive System Setup

In this section, the effectiveness of the proposed adaptive PID control scheme is evaluated through conducting a series of experiments on a prototype 1HP SPMSM drive using TMS320F28335 digital signal processor (DSP), which is extensively used at ac motor drives. Note that with the advanced development of the DSP today, the proposed scheme with the simple adaptive tuning laws does not affect much the processing time and CPU utilization to execute the algorithm compared to the traditional PID method.

Next, only two phase currents (*ia*, *ib*) are measured through the Hall sensors, and converted from analog values to digital values via 12-bit A/D converters. Note that since the stator windings are connected in star configuration, the phase-*C* current can be easily calculated from the phase-*A* and -*B* currents (*ia*, *ib*). In this paper, the space vector pulse width modulation (SVPWM) technique is used to efficiently regulate the rotor speed. Taking into account the system efficiency and control performance, the switching (or sampling) frequency and dead time are selected as 5 kHz and 2 μs, respectively.

Study Scenarios

To evaluate the effectiveness of the proposed control strategy, there are various study scenarios such as variable load torque, variable speed, parameter variations, etc. In this paper, the regulation performance of the proposed control method is evaluated by the two scenarios that include the step changes of both the load torque and desired rotor speed. Moreover, the motor parameter variations are implemented in the experiments to verify the robustness of the control system.

Actually, the electrical parameters are changed according to the temperature and stator currents during the system operation. Based on it is assumed that the stator resistance (Rs) and stator inductance (Ls) are changed as 70% and 30%, respectively, i.e., Rs = (100% + 70%) 0.43 = 0.731 and Ls = (100% 30%) 3.2 = 2.24 mH.

Also, the mechanical parameters are normally increased when the motor shaft is connected to the external mechanical load. It should be noted that some parameters such as J and B may be more heavily changed according to the operating conditions and applications. However, the proposed control system can effectively overcome these problems by using an online tuning rule which can be adapted to the variations of any system parameters. Table II summarizes the two different scenarios described above to assess the proposed algorithm.

Notice that it is not easy to directly change the motor parameters in experiments even if these values in simulations can be easily changed. As an alternative to implement the motor parameter changes in a real SPMSM drive, it can be done simply by changing the parameters in the control scheme. Therefore, in this paper, the changes of the system parameters in the controller have been made instead of changing the real system parameters in the SPMSM in order to experimentally verify the control performance of the proposed method and conventional PID method under the variations of some motor parameters (Rs, Ls, J, and B). Since the conventional PID controller possesses the control structure similar to the proposed adaptive PID controller therefore, it is also implemented for the competitive comparison.

Notice that the gains of the conventional PID controller are determined by the tuning rules of based on the pole placement technique. In this paper, the conventional PID control gains are chosen as $K1P$ = 30000, $K1I$ = 3000, $K1D$ = 100, $K2P$ = 200, and $K2I$ = 50. Note that, these values are also used as the initial values to online tune the control gains of the proposed adaptive PID controller. In (16), the positive learning-rates ($\gamma 1P$, $\gamma 1I$, $\gamma 1D$, $\gamma 2P$, and $\gamma 2I$) should be sufficiently large to guarantee a fast learning process and small time to converge. However, if they are selected to be too large, the proposed adaptive PID algorithm may become unstable.

Therefore, these values are chosen as $\gamma 1P = \gamma 1I = \gamma 1D = \gamma 2P = \gamma 2I = 0.1$ based on the fast learning process and system stability. Besides, the positive constants in the supervisory control term are chosen via extensive simulation studies as $\delta 1 = 5$ and $\delta 2 = 1$. Note that the angular acceleration obtained from is utilized in the conventional PID control scheme.

Experimental Results

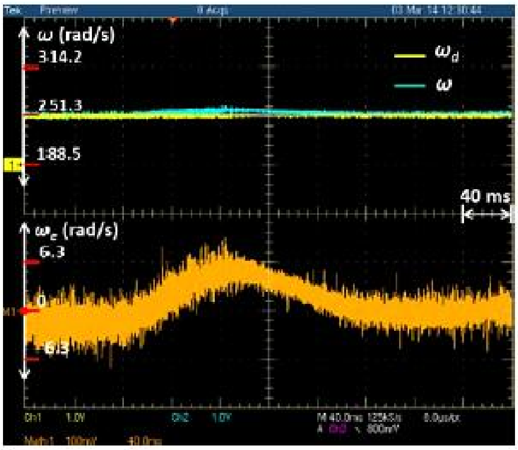
Scenario 1

A PID controller may be considered as an extreme form of a phase lead-lag compensator with one pole at the origin and the other at infinity. Similarly, its cousins, the PI and the PD controllers, can also be regarded as extreme forms of phase-lag and phase-lead compensators, respectively. It can be inferred from Fig. 3 and 4 that the regulation performance of the conventional PID control system is significantly improved after applying the adaptive tuning laws. That is, the proposed control scheme precisely tracks the desired speed with fast dynamic response (settling time: 196 ms) and small steady-state error (2.0%), under a sudden change in load torque. On the contrary, it is obvious that the conventional PID controller still shows its poor capacity when the load torque changes with a step, i.e., the settling time and steady-state error are 240 ms and 6%, respectively. It should be noted that the gains of the conventional PID controller are tuned under nominal parameters via extensive simulation studies. As shown in Fig. 4, its steady-state error is quite high because it lacks an adaptive capacity under parameter uncertainties.

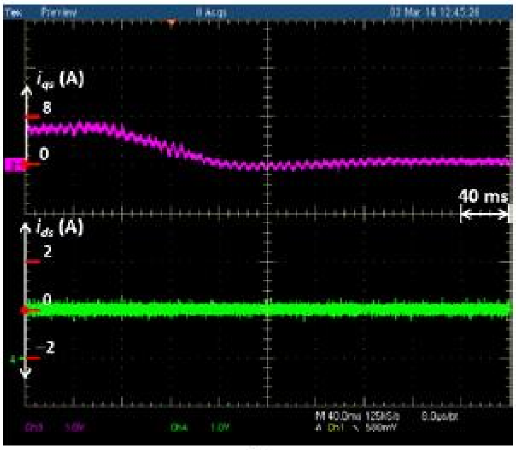
Scenario 2

In this experimental scenario, the desired speed (ωd) is suddenly changed from 125.7 rad/s to 251.3 rad/s and the load torque (TL) is set at 1 under system parameter variations. Fig. 5 and 6 show the experimental results of the proposed PID control method and the conventional PID control method. In the rotor speed can be tracked to accurately follow the desired value (steady-state error: 1.6%). On the other hand, it can be seen in that the conventional controller tracks the rotor speed with a considerable steady-state error (9.1%). In these figures, the proposed adaptive PID control scheme (settling time: 90 ms) exhibits the faster dynamic behavior than the conventional control method (settling time: 216 ms) under the parameter uncertainties the proposed adaptive PID controller can more effectively improve the control performance (i.e., faster dynamic response and smaller steady-state error) than the conventional PID controller when there exist the motor parameter variations and external load disturbances.Experimental results of the proposed adaptive PID control method when the load torque suddenly changes under system parameter variations. (a) Desired speed (ωd), rotor speed (ω), and speed error (ωe); (b) d-axis stator current (ids) and q-axis stator current (iqs).

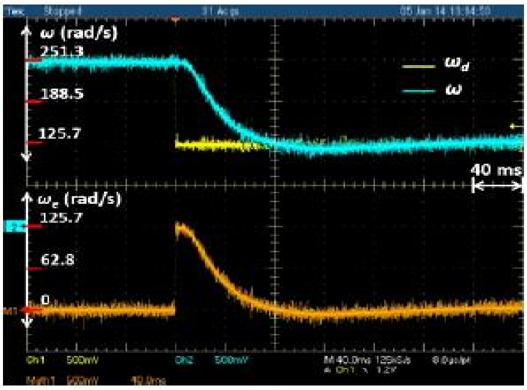
These phase voltages are compared with a high-frequency carrier signal to generate the pulsewidth-modulated (PWM) signals, which act as firing pulses for the six transistors of the inverter. Thus, these six PWM logic signals are the output of the DSP board and fed to the base drive circuit of the inverter power module. The 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.



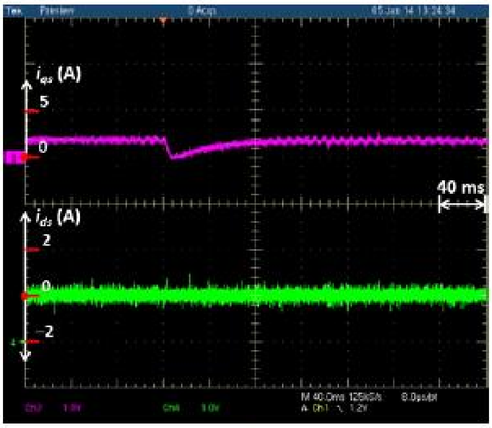
(a)



(b)



(a)



(b)

Experimental results of the conventional PID control method when the load torque suddenly changes under system parameter variations. (a) Desired speed (ωd), rotor speed (ω), and speed error (ωe); (b) d-axis stator current (ids) and q-axis stator current (iqs).

Conclusion

In this paper, an adaptive PID control method for speed control of the PMSM drives was proposed. The proposed control algorithm was simple and easy to implement in the practical applications. By using the gradient descent method, the adaptive tuning laws were proposed to auto adjust the PID gains that can achieve favorable tracking performance. The position and speed estimation errors are attenuated obviously under different operating conditions including the step load disturbance, the acceleration and the deceleration. The performance of the sensor less IPMSM drive can be improved with the proposed method. However, since the EMF-based method could not work for sensor less control at zero speed, it should combine with the saliency-based method to achieve the whole-speed-range sensor less control. Therefore, the proposed control scheme could guarantee the accurate and fast speed tracking in spite of the system parameter variations and external load disturbances.

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