論 文 16-1-6

영구자석 동기전동기의 강인한 디지털 속도 제어기법

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Robust Digital Speed Control Scheme of Permanent Magnet Synchronous Motor

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요 약

본 논문에서는 표면부착형 영구자석 동기전동기를 위하여 강인한 디지털 속도제어기를 제안한다. 제안된 속도제어기는 부하 토크관측기를 필요로 하지 않는 간단하며 부하 외란에 둔감한 디지털 제어 기법을 사용하므로 제어 성능의 저하 없이 쉽고 단순하게 구현될 수 있다. 제안된 제어 알고리즘의 성능을 검증하기 위하여, 프로토타입 표면부착형 영구자석 구동시스템을 이용하여 시뮬레이션 및 실험을 하였다. 모터 파라미터 변동 하에서 수행된 시뮬레이션 및 실험결과를 통하여 제안된 기법이 표면부착형 영구자석 동기전동기의 속도를 정확하게 제어할 수 있음을 확인하였다.

ABSTRACT

This paper proposes a robust digital speed regulator for a surface-mounted permanent magnet synchronous motor(SPMSM). The proposed speed controller uses a simple digital load disturbance resistance scheme which does not require a load torque observer, so it can be easily and simply implemented without degrading the control performance. To validate the effectiveness of the proposed control algorithm, experimental results as well as simulation results are shown under motor parameter variations using a prototype SPMSM driving system. Finally, it was confirmed that the proposed method can precisely regulate the speed of the SPMSM.

Key Words : Digital control, Permanent magnet synchronous motor, Speed control

1. Introduction

Recently, a permanent magnet synchronous motor(PMSM) has been extensively used in servo applications such as CNC machine tools, manipulators, industrial robots, and hard disk drives since it features high efficiency, low noise, and low inertia. However, the PMSM has nonlinear characteristics and uncertainties due to motor parameter variations and unknown external disturbances. Thus it is not easy for the PMSM to precisely be controlled if the linear control methods are used. To overcome these problems, some researchers have proposed various nonlinear design methods, e.g., adaptive control method[1]-[3], nonlinear feedback linearization control method[4], fuzzy control method[5]-[6]. In recent years, several papers have proposed disturbance-observer based PMSM control methods which can remarkably reject load torque variations[7]-[8]. In this case, the robust performance of the PMSM servo system cannot be guaranteed

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접수일자 : 2010. 9. 17 1차 심사 : 2010. 10. 28
심사완료 : 2010. 11. 3

under incomplete knowledge of states, PMSM model parameters, and load torque values. Furthermore, most of the previous disturbance-observer-based control design methods have developed in the continuous-time domain. In digital implementation of the previous control methods, the estimated disturbance values need to be computed via the discretized versions of the continuous-time load torque disturbance observer or other approximation methods to solve the continuous-time dynamic equations. This is very computationally burdensome.

In this paper, a simple digital control scheme is proposed which can suppress load torque disturbances without a load torque disturbance observer. First of all, the proposed speed controller is simple and it can be easily implemented without degrading the control performance. Moreover, the proposed speed regulator is insensitive to system parameter variations. It is shown that the proposed control system is asymptotically stable. Simulation and experimental results are presented under motor parameter variations to prove the effectiveness of the proposed control method using a prototype SPMSM driving system.

2. System Description

By taking the rotor coordinates of the motor as reference coordinates, a surface-mounted PMSM can be represented by the following nonlinear equation:

$$\begin{aligned} \dot{\omega} &= k_1 i_{qs} - k_2\omega - k_3 T_L \\ \dot{i}_{qs} &= -k_4 i_{qs} - k_5\omega + k_6 V_{qs} - \omega i_{ds} \\ \dot{i}_{ds} &= -k_4 i_{ds} + k_6 V_{ds} + \omega i_{qs} \end{aligned} \qquad (1)$$

where T_L represents the load torque, ω is the electrical rotor angular speed, i_{qs} is the q-axis current, V_{qs} is the q-axis voltage, i_{ds} is the d-axis current, V_{ds} is the d-axis voltage, and $k_i > 0$, $i = 1,\cdots,6$ are the parameter values given by

$$k_1 = \frac{3}{2}\frac{1}{J}\frac{p^2}{4}\lambda_m,\ k_2 = \frac{B}{J},\ k_3 = \frac{p}{2J},$$
$$k_4 = \frac{R_s}{L_s},\ k_5 = \frac{\lambda_m}{L_s},\ k_6 = \frac{1}{L_s}.$$

and p, R_s, L_s, J, B, λ_m are the number of poles, the stator resistance, the stator inductance, the rotor inertia, the viscous friction coefficient, the magnetic flux. The load torque disturbance term T_L can severely deteriorate the control performance if it is not appropriately considered. Fig. 1 shows the equivalent circuit model of the SPMSM.

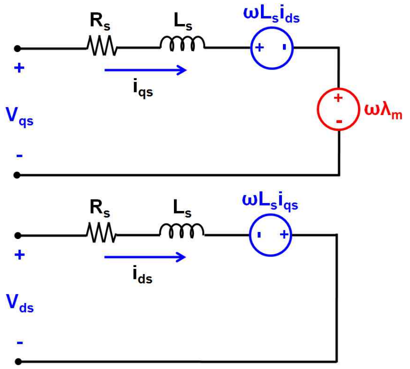


Fig. 1 SPMSM equivalent circuit

3. Digital Load Disturbance Resistance Speed Controller Design

3.1 Continuous-Time Controller Design

Let us put the following assumptions to design a robust speed regulator:

A1 : θ, i_{qs}, i_{ds} are available.
A2 : $\dot{T}_L$ can be neglected and set as $\dot{T}_L=0$.
A3 : The desired speed(ω_d) is bounded, $\dot{\omega}_d=\ddot{\omega}_d=0$.

Let us define the electrical rotor angular acceleration by $\beta = k_1 i_{qs} - k_2\omega - k_3 T_L = \dot{\omega}$. Then by using A2, (1) can be rewritten as

$$\begin{aligned} \dot{\omega} &= \beta \\ \dot{\beta} &= -k_2\beta - k_1k_4 i_{qs} - k_1k_5\omega - k_1\omega i_{ds} + k_1k_6 V_{qs} \\ \dot{i}_{ds} &= -k_4 i_{ds} + \omega i_{qs} + k_6 V_{ds} \end{aligned} \qquad (2)$$

By introducing the speed error $\omega_e = \omega - \omega_d$ and by setting V_{qs} and V_{ds} as

$$V_{qs} = \frac{1}{k_1 k_6}(u_{fq} + u_{fbq}) \quad (3)$$
$$V_{ds} = \frac{1}{k_6}(u_{fd} + u_{fbd})$$

where u_{fq} and u_{fd} are the linearizing control terms to compensate for the nonlinearities of PMSM, and u_{fbq} and u_{fbd} are the feedback control terms to stabilize the error dynamics. Let each control law be given by

$$u_{fq} = k_2\beta + k_1 k_4 i_{qs} + k_1 k_5 \omega + k_1 \omega i_{ds} \quad (4)$$
$$u_{fd} = k_4 i_{ds} - \omega i_{qs}$$
$$u_{fbq} = -K_1 \omega_e - K_2 \beta$$
$$u_{fbd} = -K_3 i_{ds}$$

where K_1, K_2 and K_3 denote positive gains. By substituting (3) and (4) into (2), the closed-loop control system can be obtained as

$$\dot{e} = \begin{bmatrix} 0 & 1 & 0 \\ -K_1 & -K_2 & 0 \\ 0 & 0 & -K_3 \end{bmatrix} e \quad (5)$$

where $e = [\omega_e, \beta, i_{ds}]^T$. The characteristic function of the error dynamics (5) is given by the following third-order polynomial

$$\left[s^2 + (K_1 + K_2)s + K_1 K_2\right] \cdot \left[s + K_3\right]$$

Therefore, the closed-loop control system is exponentially stable as long as $K_1 > 0$, $K_2 > 0$, $K_3 > 0$. Then, the following theorem can be obtained:

Theorem: Consider the closed-loop system of (1) and (4). Then, the asymptotic stability of $e = 0$ is guaranteed as long as $K_1 > 0$, $K_2 > 0$, $K_3 > 0$.

3.2 Discrete-Time Controller Design

Let us rearrange V_{qs} given in (3) as

$$V_{qs}(t) = u_{qs}(t) + u_{qd}(t) \quad (6)$$

where $u_{qs}(t)$ and $u_{qd}(t)$ are the static term and dynamic term given by

$$u_{qs}(t) = \frac{1}{k_6}(k_4 i_{qs} + k_5 \omega + \omega i_{ds}) - \frac{K_1}{k_1 k_6}(\omega - \omega_d) \quad (7)$$

$$u_{qd}(t) = \frac{(k_2 - K_2)}{k_1 k_6}\frac{d\omega}{dt} = \frac{(k_2 - K_2)}{k_1 k_6}\beta \quad (8)$$

For digital implementation under a sufficiently small sampling time T, the static terms V_{ds} and u_{qs} of V_{qs} at the sampling instant kT can be straightforwardly set as

$$u_{qs}(k) = \frac{1}{k_6}(k_4 i_{qs}(k) + k_5 \omega(k) + \omega(k) i_{ds}(k)) - \frac{K_1}{k_1 k_6}(\omega(k) - \omega_d(k)) \quad (9)$$

$$u_{qd}(k) = \frac{1}{k_6}((k_4 - K_3) i_{ds}(k) - \omega(k) i_{qs}(k)) \quad (10)$$

On the other hand, the dynamic term $u_{qd}(k)$ cannot be straightforwardly obtained. By using the relation $\beta = \dot{\omega}$ and the previous result[9],[10], $\beta(k)$ can be computed by using the following equation:

$$\beta(k) = \frac{\rho}{T+\rho}\beta(k-1) + \frac{1}{T+\rho}[\omega(k) - \omega(k-1)] \quad (11)$$

where ρ is a sufficiently small filter time constant to limit the susceptibility of the derivative term $\beta = \dot{\omega}$ to noise. Therefore, the term $u_{qd}(k)$ can be computed by the following recursive equation:

$$u_{qd}(k) = \frac{\rho}{T+\rho}u_{qd}(k-1) + \frac{(k_2 - K_2)}{k_1 k_6 (T+\rho)}[\omega(k) - \omega(k-1)] \quad (12)$$

From (9) to (12), the proposed digital control algorithm can be represented as the block diagram shown in Fig. 2.

Remark: If ρ is set as $\rho = T$, (11) becomes the Tustin's approximation. Also, if ρ is set as $\rho = 0$, then (11) becomes the backward difference approximation.

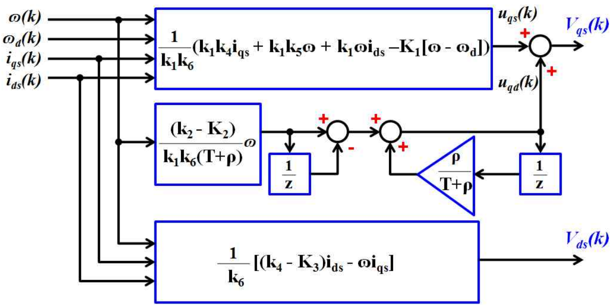


Fig. 2 Block diagram of the proposed digital control algorithm

4. Simulation and Experimental Results

For simulation and experiment, let us consider a SPMSM (1) with p = 12, R_s = 0.99[Ω], L_s = 5.82[mH], T_L = 0.5[N·m], λ_m = 0.0792[V·sec/rad], J = 0.0012[kg·m^2], B = 0.0003[N·m·sec/rad]. Fig. 3 shows the block diagram of the overall experimental control system to demonstrate the proposed control scheme. In Fig. 3, it includes a PM synchronous motor, an encoder, a brake, and a three-phase PWM inverter with a TMS320F28335 DSP controller. In this experiment, it should be noticed that all blocks in the dotted line are implemented on a Texas Instruments TMS320F28335 DSP. In both simulations and experiments, the switching frequency as well as the sampling frequency(1/T) is chosen as 5[kHz] considering the switching losses and current ripples, and a space vector PWM (SVPWM) technique is used to generate three phase voltages(V_{an}, V_{bn}, V_{cn}) applied to the motor. It is assumed that K_1 = 3061, K_2 = 3187, K_3 = 500, and ρ = 0. By referring to (9), (10), and (12), the control law can be obtained as follows:

$$\begin{aligned} u_{qs}(k) &= 0.99 i_{qs}(k) + 0.0792\omega(k) + 0.0058\omega(k) i_{ds}(k) \\ &\quad - 0.5033[\omega(k) - \omega_d(k)] \\ u_{qd}(k) &= -26.1991[\omega(k) - \omega(k-1)] \\ V_{ds}(k) &= -1.92 i_{ds}(k) - 0.0058\omega(k) i_{qs}(k) \end{aligned} \quad (13)$$

Figs. 4 and 5 show the simulation results which are performed by Matlab/Simulink about two cases: nominal parameters and 150% variations of some motor parameters(J, R_s, L_s). In this paper, the reference speed(ω_d) increases from 157.08[rad/sec] to 314.15[rad/sec] and then decreases from 314.15[rad/sec] to 157.08[rad/sec]. Fig. 4 shows the simulation results(ω, ω_d, V_{qs}, V_{ds}, i_{qs}, i_{ds}, V_{an}, i_a) about the transient response with nominal parameters. Fig. 5 shows the simulation results with 150% variations of some motor parameters(J, R_s, L_s). It is confirmed that the proposed method is very robust to system parameter variations.

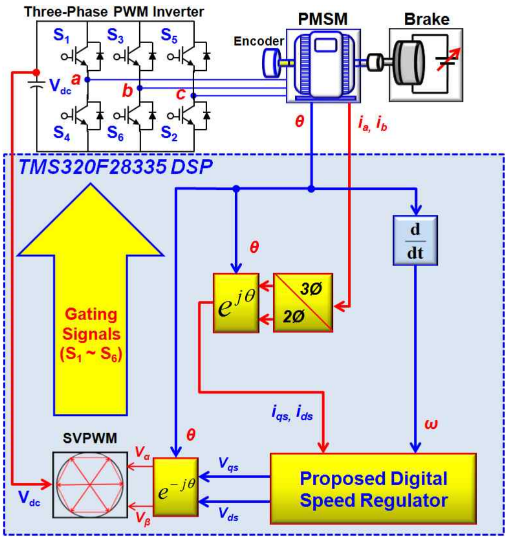


Fig. 3 Block diagram of overall experimental control system

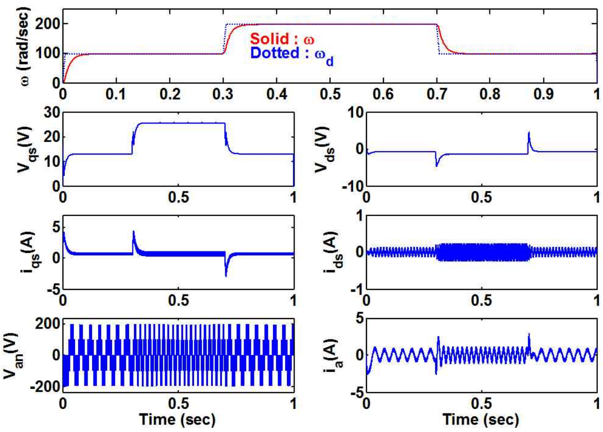


Fig. 4 Simulation results under nominal parameters

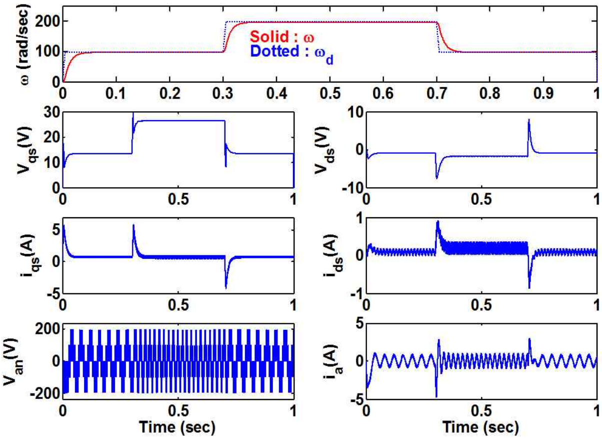


Fig. 5 Simulation results under 150% variations of some motor parameters(J, R_s, L_s)

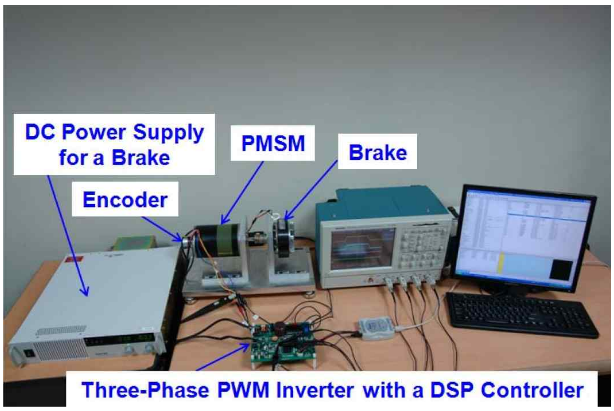
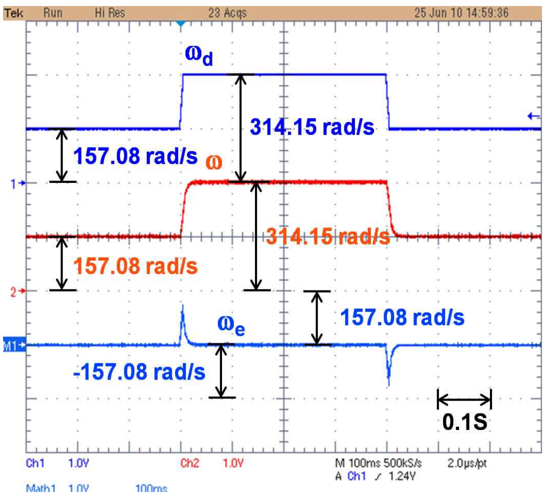
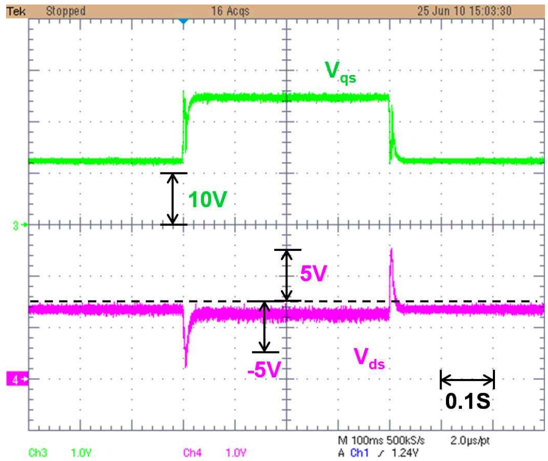


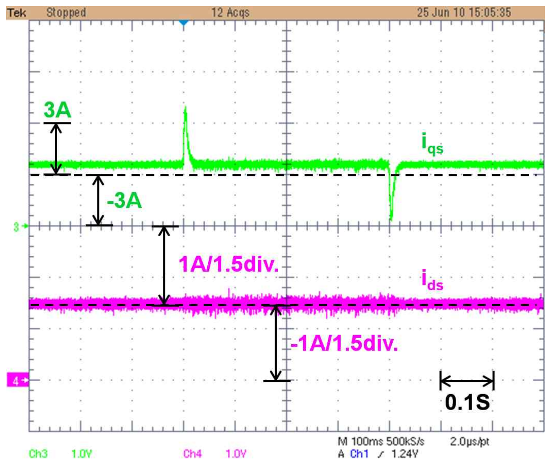
Fig. 6 Experimental test bed



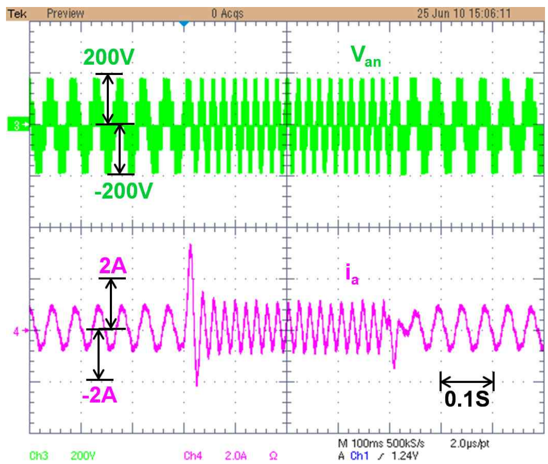
(a) Desired speed(ω_d) and measured speed(ω)



(b) q-axis voltage(V_{qs}) and d-axis voltage(V_{ds})



(c) q-axis current(i_{qs}) and d-axis current(i_{ds})



(d) Phase a voltage(V_{an}) and phase a current(i_a)

Fig. 7 Experimental results of the proposed control method

Fig. 6 shows the photograph of the testing setup used for the experimental work to demonstrate the proposed control algorithm. Fig. 7 shows the experimental results under the same condition as Fig. 4. Fig. 7(a) shows the measured speed(ω), desired speed(ω_d), and speed error(ω_e). Fig. 7(b) shows the q-axis voltage(V_{qs}) and d-axis voltage(V_{ds}). Fig. 7(c) shows the q-axis current(i_{qs}) and d-axis current(i_{ds}). Fig. 7(d) shows the phase a voltage(V_{an}) and phase a current(i_a). From simulation and experimental results, it is clearly illustrated that the proposed digital control method can be used to precisely control the speed of a surface-mounted permanent magnet synchronous motor.

5. Conclusion

In this paper, a simple digital speed control design method was proposed for a SPMSM. The proposed speed controller can be easily implemented without degrading the control performance. Moreover, it was shown that the proposed speed regulator is insensitive to motor parameter variations. From simulations and experiments, it was concluded that the proposed control system can guarantee a good performance under the system parameter variations.

이 논문은 2010년도 정부(교육과학기술부)의 재원으로 한국연구재단의 지원을 받아 수행된 기초연구사업임(2010-0009577)

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