

Velocity Control for Ripple Reduction in Permanent Magnet Synchronous Motors With Low Performance Current Sensing

YOUNGWOOL¹, DONGHOON SHIN², AND WONHEE KIM³ 

¹Department of Electrical Engineering, Chonnam National University, Gwangju 61186, South Korea

²Global R&D Center, Mando Corporation, Seongnam 13486, South Korea

³School of Energy Systems Engineering, Chung-Ang University, Seoul 06974, South Korea

Corresponding author: Wonhee Kim (whkim79@cau.ac.kr)

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ABSTRACT In this study, a velocity control method is proposed for minimizing velocity ripples in permanent-magnet synchronous motors with low-performance current sensing. The proposed method involves the use of desired currents, peak estimation, and a proposed tracking controller. The desired currents are used to achieve velocity control with only the current sensing. Peak estimation is performed to estimate the peak of the measured distorted current that occurs due to low-performance sensing. The proposed tracking controller is developed based on a proportional-integral controller using the estimated peak obtained from the measured currents. The proposed method aids in reducing velocity ripples and improving velocity tracking performance of permanent-magnet synchronous motors. The primary contribution of this method is the improvement of velocity-control performance with low-performance current sensing. The stability of the closed loop is proven by employing the input-to-state stable property. To verify the effectiveness of the proposed method, we performed simulations and experiments with different three cases.

INDEX TERMS Velocity control, ripple reduction, peak estimation, low-performance current sensing, permanent-magnet synchronous motor.

I. INTRODUCTION

Nowadays, permanent-magnet synchronous motors (PMSMs), such as electric vehicles, wind generation systems, industrial robots, air conditioners, and washing machines, are widely used in household and industrial applications owing to their high efficiency, low maintenance cost, and high power density [1]–[4]. In general, it is difficult to control a PMSM, because it is a nonlinear system with two inputs, and its control performance can be affected by variations in the parameters. Thus, a high-performance control method with a simple algorithm that exhibits fast response, high accuracy, and robustness against variations in the load torque and motor parameters is required.

Various control methods have been examined to improve the velocity control performance of PMSMs. A field programmable gate array-based velocity control method using an

adaptive fuzzy control was designed [5]. A velocity control method using predictive functional control and an extended state observer was also proposed [6]. Sliding mode controllers were proposed to compensate for the disturbances in the velocity control of PMSMs [7]–[9]. A fractional order control method was developed to control the position of a PMSM system [10]. A robust control algorithm was applied with a neural network in PMSMs to achieve velocity control [11]. A constrained state feedback speed control method was developed based on a model predictive approach [12]. A linear parameter varying H_∞ velocity control method was proposed to achieve robustness against disturbances [13]. Velocity control methods using direct torque control was designed for optimal dynamic performance [14].

Although these methods improve the velocity control performance of PMSMs, accurate information of the state variables (position, velocity, and currents) is also required for their implementation. To overcome these problems, sensorless control methods have been researched [15]–[17]. In [15],

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[16], nonlinear observers were proposed for the estimation of the full state variables using position feedback. Position sensorless control using zero-sequence carrier injection was developed [17]. High-performance current sensing is generally required to estimate the position using current feedback. For high-performance current sensing, it is important to convert the measured analog signals into digital signals using appropriate sensors. Therefore, post-digital processing using suitable analog-to-digital converters (ADCs) is important. Unfortunately, it is not always possible to use high-quality ADCs owing to cost limitations. Thus, the position information cannot be used when only current feedback with the low-cost ADC is available. Consequently, control methods with direct-quadrature (DQ) transformation cannot be used as DQ transformation requires position feedback [12].

Generally, a proportional-integral (PI) controller is adopted for velocity control of PMSMs in industrial applications owing to its simplicity, clear functionality, and effectiveness [18]–[20]. Thus, PI controllers have been widely used with low-cost ADCs that have both low sample rates and low bit resolutions to achieve velocity control. However, the use of low-cost ADCs may cause distortions during current measurements. Thus, ripples depending on the ADC sampling frequency affect the current-tracking performance. Consequently, the current ripples cause velocity ripples, thereby degrading the velocity-control performance of PMSMs.

Motivated by the aforementioned concerns, this study proposes a velocity control method to reduce the velocity ripples for PMSMs with a low-performance current sensing. This method is based on the desired currents, peak estimation, and current tracking controller. The desired currents are used to control velocity with only current sensing. Furthermore, peak estimation is applied to estimate the peak of the distorted currents that occur due to the low-performance sensing. The current tracking controller is developed based on the PI controller to improve the current-tracking performance using the estimated peak of the measured currents. This proposed controller aids in reducing the velocity ripple, and improving the velocity-tracking performance. The primary contribution of the proposed method is the improvement of the velocity-control performance using only the low-performance current sensing. Furthermore, the proposed method is enough to enable easy implementation in the PMSMs. The velocity-control performance achieved using the proposed method is validated using simulations and experiments.

II. PMSM MODELING AND PROBLEM STATEMENT

In this section, we discuss some of the limitations associated with achieving velocity control with low-performance current sensing, which is due to the use of a low-cost ADC without a position sensor for the PMSMs. PMSM modeling is first introduced in α - β coordinates, and the desired currents are subsequently derived for velocity control. Furthermore, the practical problems of the low-cost ADC for current sensing in the velocity control of PMSMs are highlighted.

A. MODELING OF PMSMs

The PMSM consists of permanent magnets mounted on the rotor surface and three-phase stator windings. The stator voltage equations of a PMSM in a stationary frame are described by [21]:

$$\begin{aligned} v_\alpha &= Ri_\alpha + L\dot{i}_\alpha - P\Phi\omega \sin(P\theta) \\ v_\beta &= Ri_\beta + L\dot{i}_\beta + P\Phi\omega \cos(P\theta) \end{aligned} \quad (1)$$

where v_α and v_β are the phase voltages [V], and i_α , and i_β are the phase currents [A]; the subscripts α and β denote the phases. In addition, θ is the rotor position [rad], ω is the rotor velocity [rad/s], R is the resistance of the phase winding [Ω], L is the inductance of the phase winding [H], Φ is the magnetic flux of the motor [Wb], and P is the number of pole pairs. The velocity dynamics are expressed as

$$\dot{\omega} = \frac{1}{J}(\tau - f_v\omega - \tau_l) \quad (2)$$

where the electromechanical torque τ is given by

$$\tau = -\frac{3}{2}P\Phi \sin(P\theta)i_\alpha + \frac{3}{2}P\Phi \cos(P\theta)i_\beta. \quad (3)$$

B. DESIRED CURRENTS FOR VELOCITY CONTROL

For the velocity control with only current sensing, the desired currents are defined as

$$i_\alpha^d = I \cos(P\theta^d), \quad i_\beta^d = I \sin(P\theta^d) \quad (4)$$

where I is a constant and θ^d is the desired position. The absolute steady-state velocity with the desired currents for a constant desired velocity, as defined in (4), is derived as shown below. For the performance analysis of the desired currents in a constant-velocity period, we assume that the currents reach the desired currents using a controller of the current-loop in a steady-state during the constant velocity period. Thus, we can assume that the position is close to the desired position. The velocity dynamics defined in (2) can be approximated as

$$\dot{\omega} = \frac{1}{J} \left[\frac{3}{2}P\Phi I(P\theta^d - P\theta) - f_v\omega - \tau_l \right]. \quad (5)$$

Using the Laplace transform, the transfer function from the desired velocity $\omega^d = \dot{\theta}^d$ and load torque τ_l to the velocity ω is given by

$$\begin{aligned} \Omega(s) &= \frac{3IP^2\Phi}{2Js^2 + 2f_v s + 3IP^2\Phi} \Omega^d(s) \\ &\quad - \frac{2s}{2Js^2 + 2f_v s + 3IP^2\Phi} T_l(s) \end{aligned} \quad (6)$$

where $\Omega(s)$, $\Omega^d(s)$, and $T_l(s)$ are the Laplace transforms of the velocity, desired velocity and load torque, respectively. If the desired velocity and load torque are step signals, then using the final value theorem, we have:

$$\lim_{t \rightarrow \infty} \omega(t) = \omega^d. \quad (7)$$

C. PROBLEM STATEMENT

ADCs are used to transfer the measured current from the sensor to the microprocessor for control algorithm implementation. Let us define the currents y_α and y_β measured by the sensing resistance in the continuous-time domain as follows:

$$y_\alpha = I_s \cos(P\theta), \quad y_\beta = I_s \sin(P\theta) \quad (8)$$

where I_s is the amplitude of the measured currents. By applying ADCs to (8), these can be expressed by the sampled data equations as follows:

$$\begin{aligned} y_{\alpha,d}(k) &= A \cos(P\omega T_{ADC}k) \\ y_{\beta,d}(k) &= A \sin(P\omega T_{ADC}k) \\ A &= \frac{2^n}{V_{ADC}} I_s, \quad \omega = 2\pi f_s, \quad f_s = \frac{1}{T_s} \end{aligned} \quad (9)$$

where A is the number of ADC bit counts expressed as an amplitude of the real measured signal, n is the ADC bit resolution, V_{ADC} is the ADC input voltage range, k and T_{ADC} are the time index number and sampling rate of ADC, respectively, and f_s is the operating frequency. The discretization process of the currents measured by the ADC is described in Fig 1. Generally, to obtain a discretized current from the ADC for an accurate velocity control, the specification of the ADC should be determined by considering the amplitude and frequency of the measured current to avoid distortion in the discretized currents. However, in many industrial applications, it is difficult to use ADCs with high specification (high sample rate/resolution) because they are expensive to manufacture. In this case, the following problems arise:

- The amplitude of the discretized current signal is distorted compared with the current measured by the ADC.
- There is a torque ripple of low frequency harmonics caused by the low ADC sample rate.

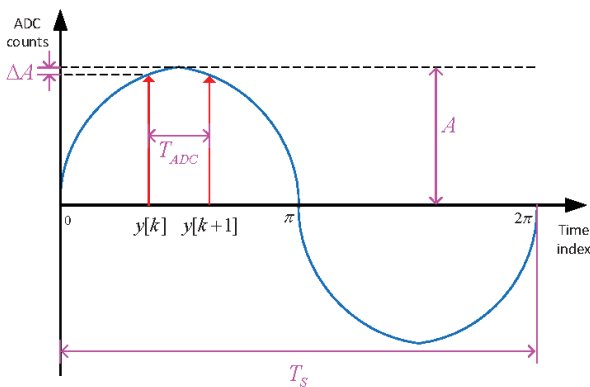


FIGURE 1. Discretization of sensing signal using ADC.

III. PEAK ESTIMATION AND CONTROLLER DESIGN

As aforementioned, it can be difficult to obtain the real peak of the measured currents owing to the limitation of the low-cost ADC specification. To resolve this problem, we propose a leak detection method for estimating the real

peak of the measured currents. Based on the estimated peak, we designed an amplitude control scheme to reduce the torque ripple and improve the velocity control performance.

A. PRELIMINARY OF NUMERICAL INTEGRAL

For the numerical integration, the Euler-Maclaurin summation formula can be used for a general analysis [22]. Furthermore, it is also useful for calculating the area of a periodic function. We consider the integral calculus of the sinusoidal function, expressed as:

$$\text{area} = \int_0^\pi M_p \sin \tau \, d\tau, \quad (10)$$

where M_p is the amplitude of the sinusoidal function and τ is the time interval of the sampling period. The area of a sinusoidal function is $\frac{2}{\pi}$. Thus, the subsequent formulas are derived using (10)

$$\begin{aligned} M_{avg} &= \frac{1}{\pi} \int_0^\pi M_p \sin \tau \, d\tau, \\ &= \frac{M_p}{\pi} (-\cos \tau)|_0^\pi, \\ &= \frac{2}{\pi} M_p, \end{aligned} \quad (11)$$

where M_{avg} is the average value of the positive half-phase of sinusoidal function ($0 - \pi$).

B. PEAK ESTIMATION USING THE AREA INTEGRAL

In the previous subsection, we discussed the relationship between the average and peak values of a period sinusoidal function. Similarly, in this subsection, we introduce the design steps for the peak estimation of sensing currents.

■ Peak Estimation Design Procedure

Let us consider the measured currents with offset bias from (9) as follows

$$\begin{aligned} y_{\alpha,d}(k) &= A \cos(P\omega T_{ADC}k) + b_c, \\ y_{\beta,d}(k) &= A \sin(P\omega T_{ADC}k) + b_s \end{aligned} \quad (12)$$

where b_c and b_s denote the offset values of each phase, respectively.

- **Step 1: Decision regarding the number of sampled sensing currents (L)**

The index number of the measured current signals is set by $k = 0, 1, 2, \dots, L - 1$.

Considering the fixed period T_s of the sensing currents by this assumption, and applying N periods of sensing currents, the total number of sampled data is approximated by $L = \text{floor}(NT_s/T_{ADC})$. To obtain a uniformly distributed peak, N is set by considering the acceptable opportunity between the accuracy and control time delay caused by the unnecessarily large N results in the slow control period.

- **Step 2: Offset Estimation (b)**

Various calibration methods for non-ideal signal components have been studied for the estimation of the measured current-signal offsets [23], [24]. In this step,

the desired frequency is set by a rational number to obtain uniformly distributed measured signals. The offset degrades the peak estimation performance. The offset calibration should be considered when summing the sensing current.

► Offset estimation:

$$\begin{aligned}\hat{b}_c &= \frac{1}{L} \sum_{k=0}^{L-1} y_{\alpha,d}(k) = \frac{1}{L} \sum_{k=0}^{L-1} A \cos(P\omega T_{ADC}k) + \frac{b_c}{L} \\ \hat{b}_s &= \frac{1}{L} \sum_{k=0}^{L-1} y_{\beta,d}(k) = \frac{1}{L} \sum_{k=0}^{L-1} A \sin(P\omega T_{ADC}k) + \frac{b_s}{L} \\ \hat{b} &= \frac{\hat{b}_c + \hat{b}_s}{2}\end{aligned}\quad (13)$$

where b_c and b_s are the offsets of phases α and β , respectively.

• Step 3: Peak estimation and offset calibration

Without a loss of generality, we directly calculate the peak of the sensing currents with the offset calibration. The absolute value of the measured current is required for peak estimation in the negative half-phase of the sinusoidal function (sensing currents). From (11), we have the following peak estimation law.

► Peak Estimation with offset calibration:

$$\begin{aligned}\hat{M}_{p,c} &= \frac{\pi}{2} \frac{1}{L} \sum_{k=0}^{L-1} |y_{\alpha,d}(k) - \hat{b}_c| \\ &= \frac{\pi}{2} \frac{1}{L} \sum_{k=0}^{L-1} |A \cos(P\omega T_{ADC}k)| \\ \hat{M}_{p,s} &= \frac{\pi}{2} \frac{1}{L} \sum_{k=0}^{L-1} |y_{\beta,d}(k) - \hat{b}_s| \\ &= \frac{\pi}{2} \frac{1}{L} \sum_{k=0}^{L-1} |A \sin(P\omega T_{ADC}k)|.\end{aligned}\quad (14)$$

• Step 4: Estimated currents using estimated peaks

From above Step 3, we have the calibration errors between an amplitude of the measured currents and estimated peak as follows:

$$e_{M\alpha} = A - \hat{M}_{p,c}, \quad e_{M\beta} = A - \hat{M}_{p,s} \quad (15)$$

Here, since the amplitude of the measured currents has a finite extremal value, the following assumption is reasonable.

Assumption 1: Because $\hat{M}_{p,c}$, $\hat{M}_{p,s}$, and A have a finite amplitude, $e_{M\alpha}$ and $e_{M\beta}$ are bounded as

$$e_{M\alpha} \leq \epsilon_\alpha, \quad e_{M\beta} \leq \epsilon_\beta \quad (16)$$

Then let us define estimated currents using the estimated peak of measured currents as follows:

$$\begin{aligned}\hat{y}_{\alpha,d}(k) &= \hat{M}_{p,c} \cos(P\omega T_{ADC}k) \\ \hat{y}_{\beta,d}(k) &= \hat{M}_{p,s} \sin(P\omega T_{ADC}k).\end{aligned}\quad (17)$$

Therefore, we can finally obtain the estimated currents in the continuous-time domain from (9) as follows:

$$\begin{aligned}\hat{y}_\alpha &= I_{s,\alpha} \cos(P\theta) \\ \hat{y}_\beta &= I_{s,\beta} \sin(P\theta)\end{aligned}\quad (18)$$

where $I_{s,\alpha} = \frac{V_{ADC}}{2^n} M_{p,c}$ and $I_{s,\beta} = \frac{V_{ADC}}{2^n} M_{p,s}$ are the amplitudes of the estimated currents $\hat{y}_\alpha$ and $\hat{y}_\beta$, respectively.

C. CONTROLLER DESIGN AND STABILITY ANALYSIS

The controller design using estimated currents based on the peak estimation algorithm is discussed herein. First, let us define current tracking errors as follows:

$$\begin{aligned}e_\alpha &= i_\alpha^d - y_\alpha = i_\alpha^d - i_\alpha \\ e_\beta &= i_\beta^d - y_\beta = i_\beta^d - i_\beta.\end{aligned}\quad (19)$$

Note that, y_α and y_β are respectively equal to i_α and i_β by definition of the measured currents in section II.C. Then the estimated tracking errors and the integration terms of estimated tracking errors are defined as

$$\begin{aligned}\hat{e}_{z\alpha} &= \int_0^t \hat{e}_\alpha d\tau \\ \hat{e}_{z\beta} &= \int_0^t \hat{e}_\beta d\tau \\ \hat{e}_\alpha &= i_\alpha^d - y_\alpha = i_\alpha^d - \hat{i}_\alpha \\ \hat{e}_\beta &= i_\beta^d - y_\beta = i_\beta^d - \hat{i}_\beta.\end{aligned}\quad (20)$$

Theorem 1: Given voltage equation of PMSM (1), suppose that proposed tracking controllers are designed as

$$\begin{aligned}v_\alpha &= R\hat{i}_\alpha + L\dot{i}_\alpha^d + k_P\hat{e}_\alpha + k_I\hat{e}_{z\alpha} \\ v_\beta &= R\hat{i}_\beta + L\dot{i}_\beta^d + k_P\hat{e}_\beta + k_I\hat{e}_{z\beta}\end{aligned}\quad (21)$$

where k_P and k_I are the controller gains and positive constant. Then, $\hat{e}_\alpha$ and $\hat{e}_\beta$ are input-to-state stable.

Proof: From (1) and (21), tracking error dynamics are given by

$$\dot{\hat{e}} = A\hat{e} + Bd \quad (22)$$

where

$$\hat{e} = \begin{bmatrix} \hat{e}_{z\alpha} \\ \hat{e}_\alpha \\ \hat{e}_{z\beta} \\ \hat{e}_\beta \end{bmatrix}, \quad A = \begin{bmatrix} 0 & 1 & 0 & 0 \\ -\frac{k_I}{L} & -\frac{k_P}{L} & 0 & 0 \\ 0 & 0 & 0 & 1 \\ 0 & 0 & -\frac{k_I}{L} & -\frac{k_P}{L} \end{bmatrix},$$

$$B = \begin{bmatrix} 0 & 0 & 0 & 0 \\ \frac{R}{L} & -\frac{P\Phi}{L} & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & \frac{R}{L} & \frac{P\Phi}{L} \end{bmatrix}, \quad d = \begin{bmatrix} d_1 \\ d_2 \\ d_3 \\ d_4 \end{bmatrix} = \begin{bmatrix} \tilde{i}_\alpha \\ \omega \sin(P\theta) \\ \tilde{i}_\beta \\ \omega \cos(P\theta) \end{bmatrix}$$

and $\tilde{i}_\alpha = i_\alpha - \hat{i}_\alpha$ and $\tilde{i}_\beta = i_\beta - \hat{i}_\beta$ are estimation errors of i_α and i_β , respectively. By definition of (4), we have following assumption

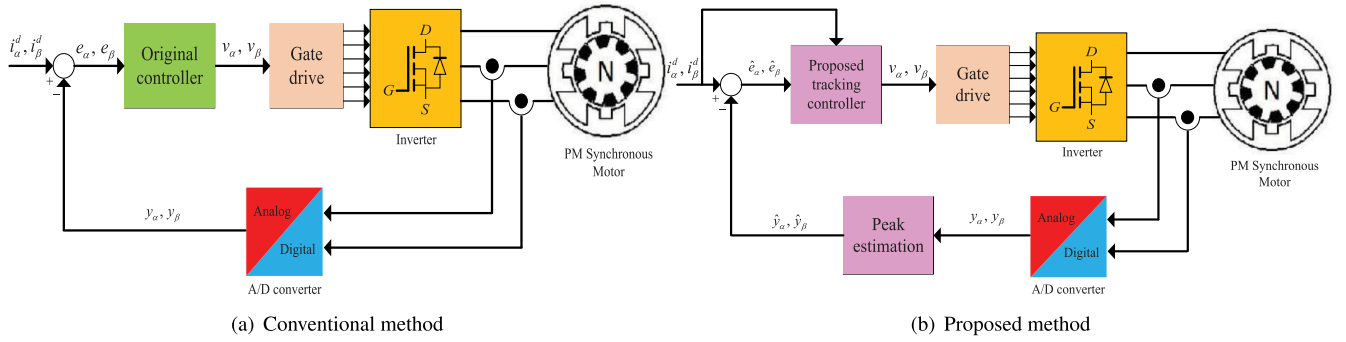


FIGURE 2. Block diagrams of the conventional and proposed method with the low-cost ADC.

Assumption 2: For constant desired velocity, actual velocity ω is bounded as

$$|\omega| < \omega_{\max}. \quad (23)$$

Also, we have following inequality for input matrix d

$$\begin{aligned} d_1 &= i_\alpha - \hat{i}_\alpha \\ &= (I_s - I_{s,\alpha}) \cos(P\theta) \leq |I_s - I_{s,\alpha}| < \gamma_\alpha \\ d_2 &= \omega \sin(P\theta) \leq |\omega \sin(P\theta)| \leq \omega < \omega_{\max} \\ d_3 &= i_\beta - \hat{i}_\beta \\ &= (I_s - I_{s,\beta}) \cos(P\theta) \leq |I_s - I_{s,\beta}| < \gamma_\beta \\ d_4 &= \omega \cos(P\theta) \leq |\omega \cos(P\theta)| \leq \omega < \omega_{\max} \end{aligned} \quad (24)$$

where $\gamma_\alpha, \gamma_\beta$, and $\omega_{\max}$ are positive and bounded real number. Therefore, matrix A is always Hurwitz for the controller gains $k_I > 0$ and $k_P > 0$, and B and d are bounded matrices. Then it is concluded that the $\hat{e}_\alpha$ and $\hat{e}_\beta$ are input-to-state stable [25].

Block diagrams of the conventional and proposed methods using the low-cost ADC are shown in Fig. 2. Note that the both methods do not require position sensing. It uses only current sensing along with the low-cost ADC.

IV. SIMULATIONS AND EXPERIMENTS

To validate the effectiveness of the proposed method, we performed simulations and experiments by comparison with the conventional method which has a same control structure in (21) with original measured currents ($i_\alpha = y_\alpha$ and $i_\beta = y_\beta$). Two different types of ADCs are defined as follows:

[High cost ADC:] $n = 14$ bits, $T_{ADC} = 10$ kilosamples per second (KSPS)

[Low cost ADC:] $n = 8$ bits, $T_{ADC} = 2$ KSPS.

Three cases were tested to evaluate the effect of the proposed method:

[Case 1:] Using the conventional method using a high-cost ADC

[Case 2:] Using the conventional method using a low-cost ADC

[Case 3:] Using the proposed method using low cost ADC.

The parameters of the PMSM, modulation gains used in case 3, and controller gains are listed in Table 1. The utilized velocity and current reference profiles are shown in Fig. 3.

TABLE 1. PMSM parameters, modulation and control gains.

Plant parameters and modulation gains		
Plant parameters	L_r	0.66 mH
	R_r	0.251 Ω
	J_r	3.24×10^{-5} kg·m ²
	Φ_r	16.8×10^{-3} Wb
	P_r	4
	f_v	2×10^{-3} N·m·s/rad
Modulation gains	L	1000, 1500, 2000
Control gains	k_P	50
	k_I	10

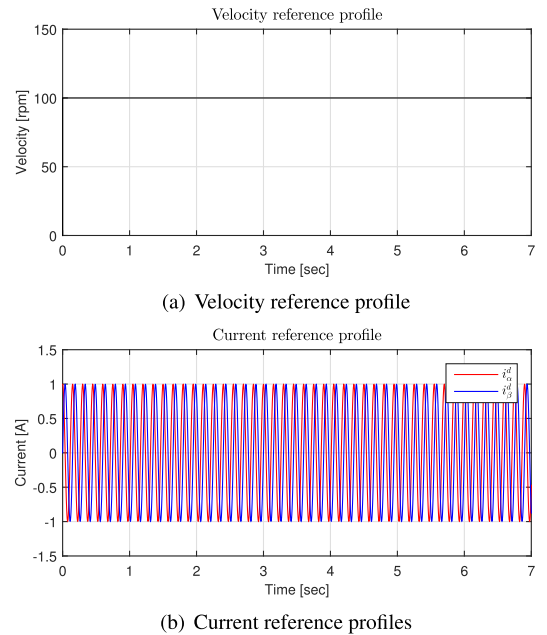


FIGURE 3. Velocity and current reference profiles.

A. SIMULATION RESULTS

The simulation was carried out using Matlab/Simulink. The PMSM model in the sim.power toolbox of Simulink was

used. The measured currents of cases 1 and 2 are shown in Fig. 4(a) and 4(b), respectively. Case 1 has a uniform peak compared with that of case 2, owing to the high sampling rate and bit resolution of the high cost ADC. Fig. 4(c) shows the frequency spectra of the measured currents for cases 1 and 2. For both cases, the fundamental frequency $P\omega$ was observed as the dominant frequency. However, there were 1st and 2nd side-band harmonics in case 2 only. Based on the discussions in the previous sections, these harmonics directly affect the torque ripples. The peak estimation performances of the proposed method are shown in Fig. 5. To optimize the estimation performance, the modulation gains should be carefully determined because a trade-off relationship exists between fast settling and transient oscillations. Further, increasing L leads to a peak phenomenon in the transient oscillations and an estimated ripple reduction in the steady state. Fig. 6 shows the velocity tracking errors of the steady-state responses of the three cases. From Figs. 6(a) and 6(b), the velocity ripples depend on the resolution and sample rate of the ADC. Thus, ripples occurred in case 2 owing to the low-cost ADC. When the proposed method was applied, the ripples caused by the low-cost ADC were considerably reduced in case 3. Despite

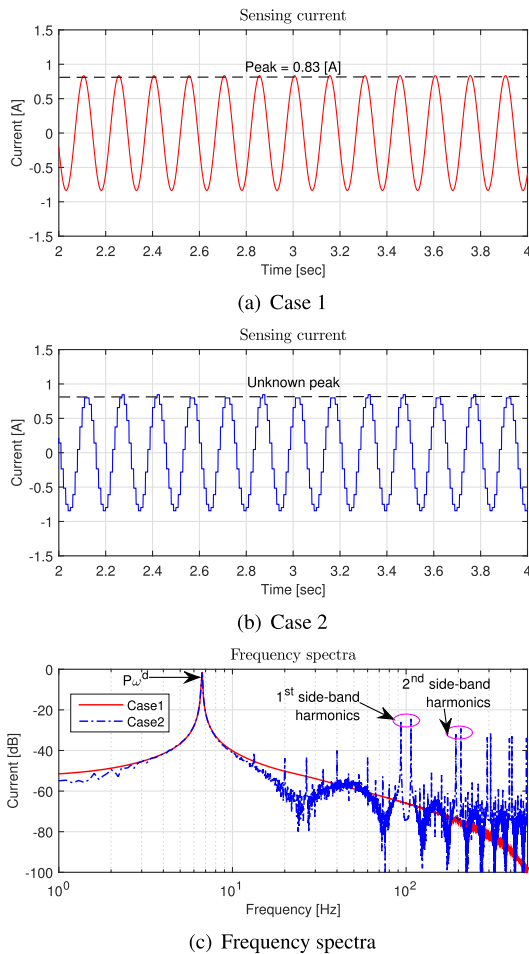


FIGURE 4. Measured currents and frequency spectra of case 1 and 2.

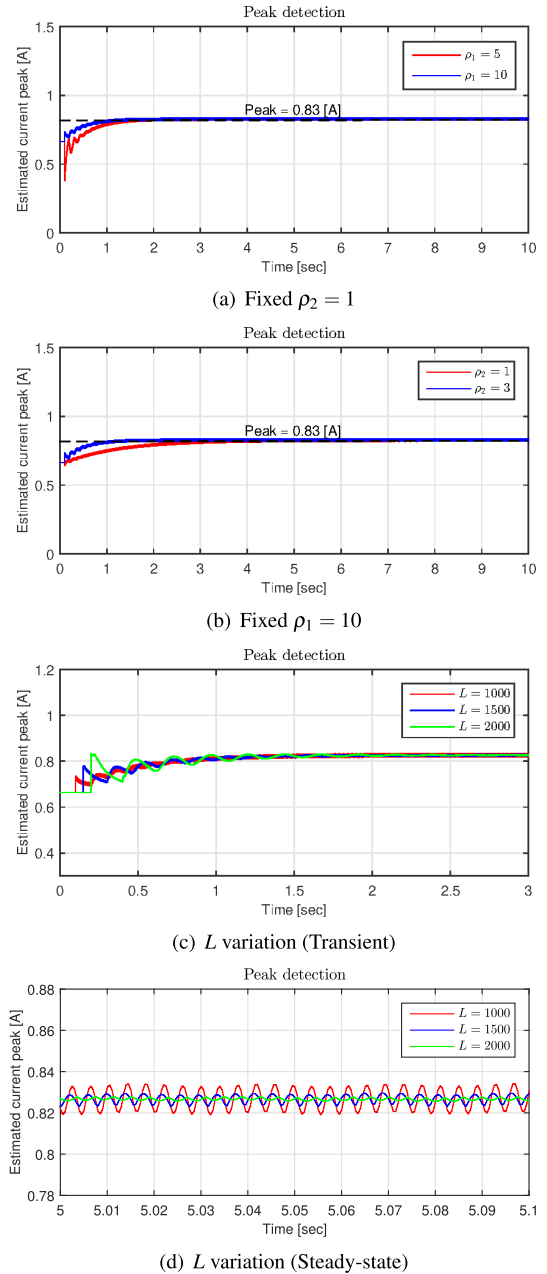


FIGURE 5. Peak estimation performance of the proposed method: (a) ρ_1 variation, (b) ρ_2 variation, (c) L variation (zoom-in along the x-axis), and (d) L variation (zoom-in along the y-axis).

using the ADC with a lower performance than that of case 1, the velocity tracking performance of case 3 was similar to that of case 1.

B. EXPERIMENT RESULTS

Figure 7 shows the experimental setup for the implementation of the proposed method. The experiments were performed using ControlDesk software, two RapidPro units, and a MicroAutoBox embedded computer (dSPACE). To apply ControlDesk, we generated a real-time control code in Matlab/Simulink. MicroAutoBox supported the 40 kHz sample

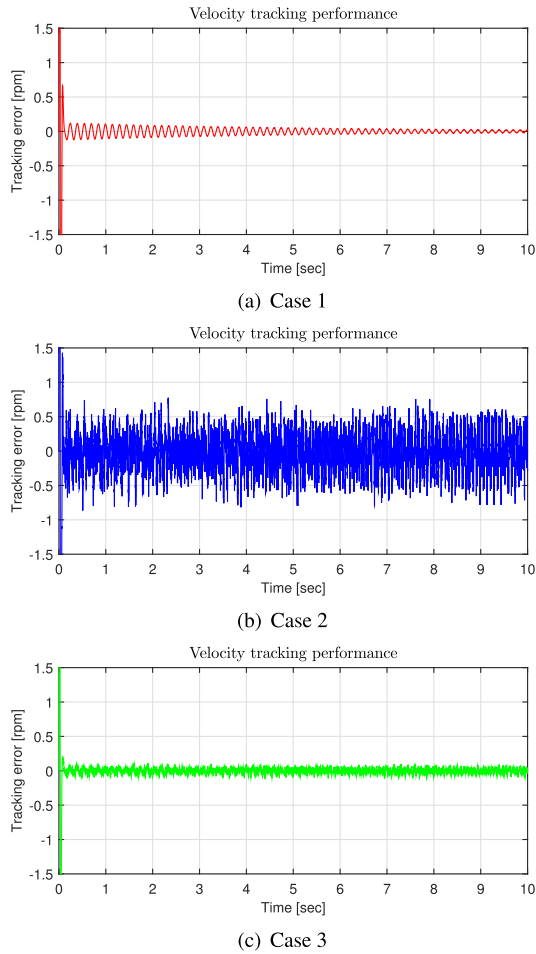


FIGURE 6. Velocity tracking errors of the three cases in steady-state responses.

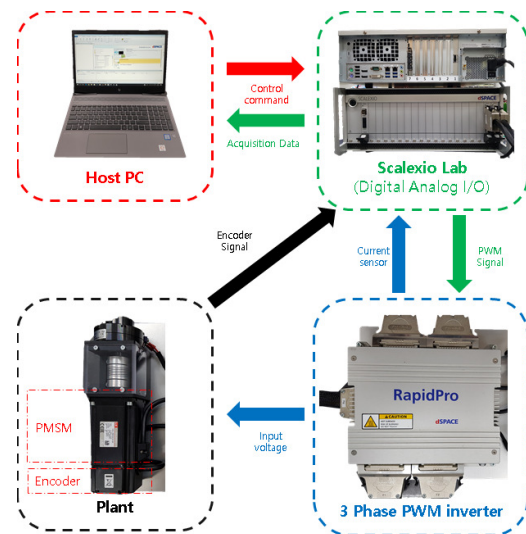


FIGURE 7. Experimental setup.

rate. The pulse width modulation switching frequency was 20 kHz. Each RapidPro unit had three half bridge power stage modules, and the two switches of each half bridge were driven by complementary signals with some dead time to

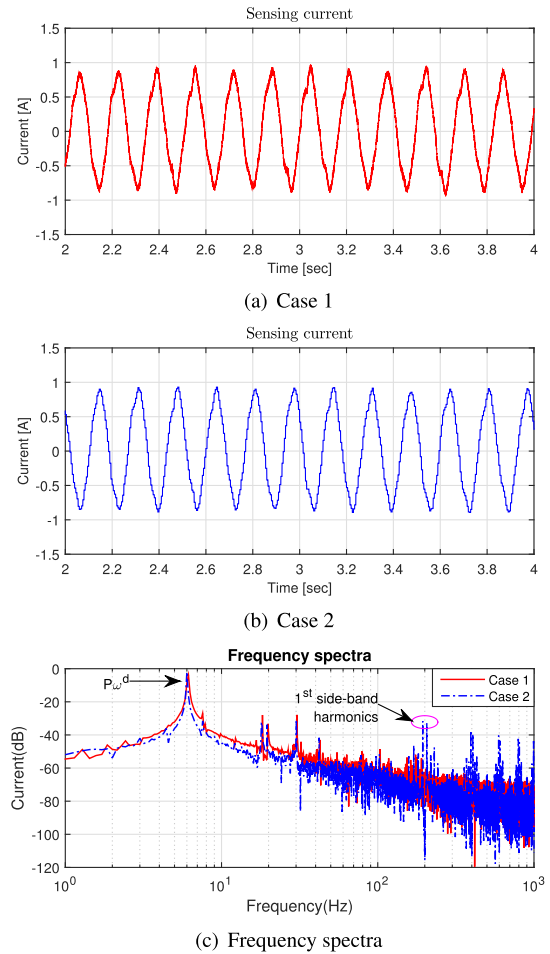


FIGURE 8. Measured currents and frequency spectra of case 1 and 2.

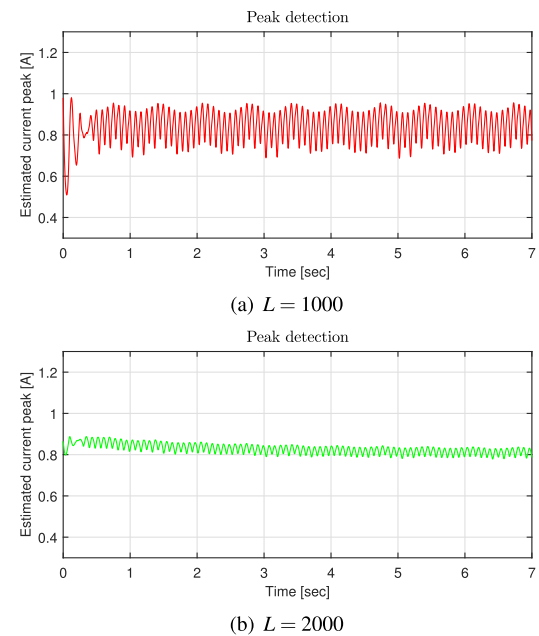


FIGURE 9. Peak estimation performances of case 3 (L variation).

avoid feed-through faults. A personal computer was used to implement and compile the control program in C-language, and to upload the executable output code to MicroAutoBox.

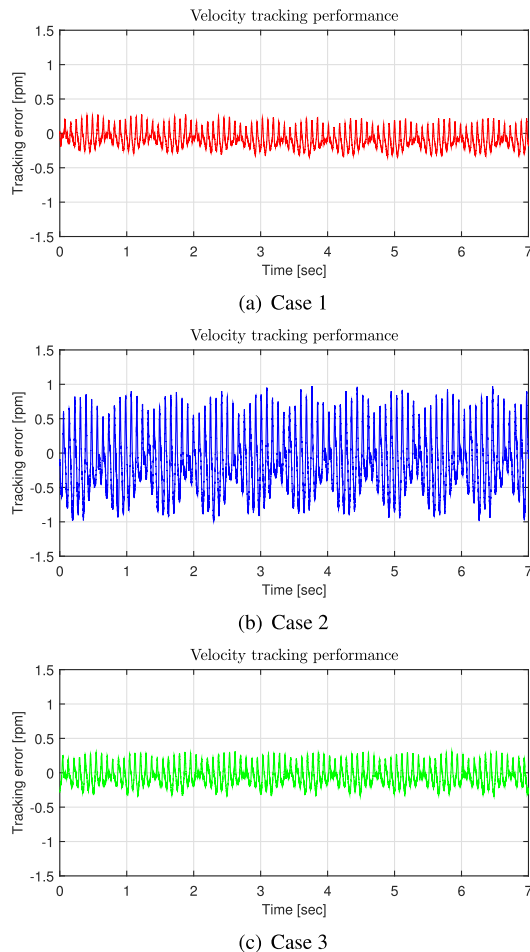


FIGURE 10. Velocity tracking errors of the three cases in steady-state responses.

The measured currents of cases 1 and 2 are shown in Fig. 8. In the simulation results, there was a distortion of the measured current due to the low specification of the low-cost ADC in case 2. Figure 8(c) shows the side-band harmonics of the ADC sampling frequency for case 2. Unlike the simulation results, low frequency harmonics below 50 Hz were observed. These low frequency components resulted from the influence of the mechanical couplers. The peak estimation performances of the proposed method are shown in Fig. 9. The ripple bands of the estimated peak changed with the L values. Although large L values lead to a reduction in the ripples of an estimated peak, they can degrade the settling performance of peak estimation. The velocity tracking performances of the three cases are shown in Fig. 10. Ripples were observed in case 1 for a low resolution and low sample rate. In the simulation results, we can see that the ripples of case 1 are larger than those of case 2. Moreover, the ripples caused by the low-cost ADC were reduced by the proposed method in case 3. Furthermore, even though a low-cost ADC was used, the ripples in case 3 were relatively small compared with those in case 1. The velocity ripples in case 3 were considerably smaller than those of case 2 because of the effect of the proposed tracking controller.

V. CONCLUSIONS

The proposed method was developed in this study to reduce velocity ripples with a low-cost ADC to achieve improved velocity control performance of PMSMs. The proposed method involves the use of the desired currents, peak estimation, and the proposed tracking controller. The desired currents were used to achieve velocity control by current sensing only. Peak estimation was performed to estimate the peak of the distorted currents due to the low-performance sensing. The proposed tracking controller was developed based on the PI controller using the estimated peak of the measured currents. The stability of the closed loop was proven by employing the input-to-state stable property. The performance of the proposed method was validated through simulations and experiments. It was observed that this method reduced the velocity ripples and improved the velocity tracking performance with low-performance current sensing.

FUTURE WORKS

Solving practical problem for state constraints is important to enhance tracking performance. Our possible future research direction is to design an optimization method considering physical state constraints such as ADC bit resolution, input voltage range.

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YOUNGWOOL LEE received the B.S. degree in electrical engineering from Chungbuk National University, Cheongju, South Korea, in 2010, and the Ph.D. degree in electrical engineering from Hanyang University, Seoul, South Korea, in 2017. From 2017 to 2018, he was a Research Scientist with the Ulsan National Institute of Science Technology, and a Visiting Scholar with the Department of Mechanical Engineering, University of California at Berkeley, Berkeley, CA, USA. From 2018 to 2019, he was with the Memory Division, Samsung Electronics Ltd., South Korea. From 2019 to 2020, he was an Assistant Professor with the Department of Electronics Engineering, Pai Chai University, Daejeon, South Korea. In 2020, he joined the Faculty of Chonnam National University, Gwangju, South Korea. His main research interests include nonlinear and optimal controller design, and motion control.



DONGHOON SHIN received the B.S. degree in electrical and computer engineering and the Ph.D. degree in electrical engineering from Hanyang University, Seoul, South Korea, in 2009 and 2016, respectively. He is currently a Senior Research Engineer with the Global R&D Center, Mando Corporation, South Korea. His main research interests include nonlinear control and observer as well as their industrial applications.



WONHEE KIM received the B.S. and M.S. degrees in electrical and computer engineering and the Ph.D. degree in electrical engineering from Hanyang University, Seoul, South Korea, in 2003, 2005, and 2012, respectively. From 2005 to 2007, he was with Samsung Electronics Company, Suwon, South Korea. In 2012, he was with the Power and Industrial Systems Research and Development Center, Hyosung Corporation, Seoul. In 2013, he was a Postdoctoral Researcher

with the Institute of Nano Science and Technology, Hanyang University, and a Visiting Scholar with the Department of Mechanical Engineering, University of California at Berkeley, Berkeley, CA, USA. From 2014 to 2016, he was with the Department of Electrical Engineering, Dong-A University, Busan, South Korea. He is currently an Associate Professor with the School of Energy Systems Engineering, Chung-Ang University, Seoul. His current research interests include nonlinear control and nonlinear observers, as well as their industrial applications.

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