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Experimental Modeling and Direct Digital Control of PMSM

Taesu Kang[1], Min-Seok Kim[2], Kiyong Lee[1], Sa Young Lee[3], Young Chol Kim[1*]
*[1]Dept. of Electrical Eng. & Computer Science, Chungbuk National Univ. 1, Chungdae-ro, Seowon-gu, Cheongju, 362-763, Korea, (*corresponding author: yckim@cbu.ac.kr)*
[2] Central Research Institute, Korea Hydro & Nuclear Power Co., Ltd, 70, Yuseong-daero 1312beon-gil, Yuseong-gu, Daejeon, Korea
[3] Dept. of Electrical Eng., Myongji College, 134, Gajwa-ro, Seodaemun-gu, Seoul, Korea

Abstract

This paper deals with the input-output modeling of a vector controlled PMSM drive system and the design of a direct digital controller for speed control. We present a discrete-time modeling technique that allows one to identify an equivalent model experimentally from the controller output to the controller input while the system is operating in closed loop. Once a discrete-time model is obtained, a digital controllers are directly designed so that it satisfies the given time response specifications. Modeling and controller design are performed separately for current control and speed control loops. The proposed approach has been demonstrated by experiments. The experimental set up consists of a surface mounted PMSM (5 KW, 220V, 8 poles) equipped with a flywheel load of 220Kg and a digital controller using DSP (TMS320F28335). It is shown that all the experimental responses coincide closely with those of target model.

Keywords: Permanent magnet synchronous motor (PMSM), input-output modelling, closed-loop identification, discrete-time characteristic ratio assignment (DCRA)

1 Introduction

Permanent magnet synchronous motor (PMSM) is widely used in the manufacturing industry, robot systems, and in many other applications because of its excellent features such as high efficiency, low cost, minimal maintenance, and high power density. To achieve the good performance for the cases that load and speed command vary, various design methods for PMSM speed control have been developed [1, 2, and wherein references]. In most non-adaptive methods, all the parameters of the PMSM model must be previously obtained in order to design a controller analytically. However, it is not easy to identify certain parameters (e.g., inertia and friction coefficient) precisely. Main drawbacks of adaptive schemes are that the adaptation algorithm may be very sensitive to the outliers of the measured data and impossible to retune the controller gains because of the complexity of adaptive controller.

The adjustable speed drive system of the vector controlled PMSM consists of generally two control loops: the current loop and the speed loop. To design a good controller for the PMSM drive system, it is necessary to have a precise model for each control loop. The modeling requires overcoming some difficulties caused by variations of the inertia moment, nonlinear characteristics at different speed levels and at different loads.

This article deals with the input-output modeling of a vector controlled PMSM drive system and the design of a direct digital controller for speed control. We present a modeling technique that allows one to model all the elements from the controller output to the controller input as a linear model while the system is operating in closed loop. We first obtain a discrete-time model for current loop, a digital current controllers are directly designed using the discrete-time characteristic ratio assignment (DCRA) so that it satisfies the given time response specifications. Next all the elements that are included in the speed control loop can be regarded as another plant. Similar to the current loop, a discrete-time model for the speed control loop is identified using the closed-loop identification technique [3, 4]. Then a digital speed controller is designed using the DCRA [5-7].

The proposed approach has been demonstrated by experiments. The experimental setup consists of a surface mounted PMSM (5 KW, 3ϕ 220V, 8 poles, Higen) equipped with a flywheel load of 220Kg and a digital controller using DSP (TMS320F28335). It is shown that all the experimental responses coincide closely with those of reference model.

2 Experimental modeling of PMSM

2.1 PMSM control system in dq frame and formulation of modeling

The mathematical model of a surface mounted PMSM is represented in the rotational two-phase frame as follows:

$$v_d = R_s i_d + L_d \frac{di_d}{dt} - L_q \omega_r i_q \, , \tag{1}$$

$$v_q = R_s i_q + L_q \frac{di_q}{dt} + L_d \omega_r i_d + \omega_r \psi_m \, , \tag{2}$$

$$J \frac{d\omega_r}{dt} = K_t i_q - T_L - B\omega_r \, , \tag{3}$$

where i_d and i_q are the d axis and q axis stator currents, and v_d and v_q for the d axis and q axis stator voltages, respectively. Resistance and inductance of the stator coil are denoted by R_s and L_d (L_q) . ω_r is the rotor angular velocity, T_L is the load torque. ψ_m is the flux linkage. J, K_t, and B are the rotor inertia, torque constant, viscous friction coefficient, respectively.

We consider here that the design procedure is based on the vector control framework, as shown in Figure 1. Two kinds of cascade controllers are included in a speed tracking loop and two current tracking loops.

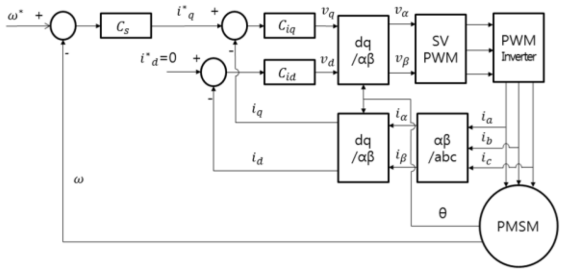

Figure 1: Block diagram of the PMSM control system

The currents i_α and i_β are obtained by taking the Clarke transformation of three phase line currents of the PMSM, and then the Park's transformation of i_α and i_β yields i_d and i_q, wherein $\theta = \omega_r t$.

The advantage of using Park's transformation is explained exactly by the fact that sinusoidal signals with angular frequency ω_r are seen as constant signals in the dq reference frame. Thus it is possible for a simple controller like PI controller to achieve good tracking performance in the current loops.

In this section, we concentrate on the experimental modeling of equivalent plants. The modeling consists of two stages. Note that the dynamics of current loop is not affected by the speed controller. This implies that the current loop can be modelled independently. All the components connected from v_q to i_q in the current loop are characterised as a discrete-time linear transfer function model at an operating condition. The system between v_d and i_d can be identified by another model similarly. Based on these models, the digital controllers for the current tracking are designed, as will be seen in Section 3. Once the current controllers are designed, the second stage is to identify the equivalent plant from i_q^* to ω including the current controllers. Then the design of digital speed controller is carried out from this model. The closed loop identification method is used for the modeling. The details are represented in next subsections.

2.2 Experimental modeling of the plant in the current loop

The mathematical model of a surfaced mounted PMSM is represented in a nonlinear equation as shown in (1)~(3). The reference current, i_d^*, is usually set as a zero, i.e., $i_d^* = 0$. Supposing that an operating condition be defined as a constant load and a constant rotor speed, and neglecting the interconnection between d and q axes, the q axis current loop can be expressed in a single loop system including a discrete-time linear time-invariant (LTI) plant, as shown in Figure 2. Wherein G_c denotes the equivalent plant model from v_q to i_q and C_c for a cascade current controller to be designed.

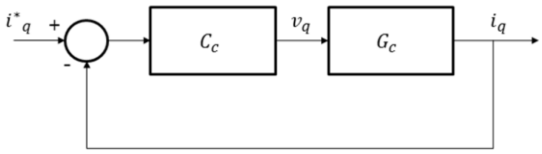

Figure 2 An equivalent model of the q-axis current loop

The problem here is to identify the model G_c experimentally. It is necessary for the model identification to be carried out in the closed loop because the PWM inverter is included in the system from v_q to i_q and various operating conditions may also be considered. The closed-loop output error (CLOE) method [3, 4] will be used for this purpose.

An identified model of the discrete-time plant G_c is defined as

$$\hat{G}_C(z^{-1}) := \frac{i_q}{v_q} = \frac{\hat{B}_C(z^{-1})}{\hat{A}_C(z^{-1})}, \tag{4}$$

where

$$\hat{A}_c(z^{-1}) = 1 + \hat{a}_{c1} z^{-1} + \cdots + \hat{a}_{nac} z^{-nac},$$

$$\hat{B}_c(z^{-1}) = \hat{b}_{c1} z^{-1} + \cdots + \hat{b}_{nbc} z^{-nbc}.$$

To apply the CLOE to the q axis current loop, we first select an arbitrary current controller $C_c = C_{iq}$ provided that the closed loop is stable. If we consider the so called R-S-T configuration instead of cascade structure C_c, the CLOE identification can be represented by the block diagram in Figure 3. In the Figure 3, r_o is equal to the q axis reference current i_q^* and r_t is the external excitation superposed onto r_o for the purpose of identification. This test input is usually chosen by a pseudo random binary sequence (PRBS) [4]. The R-S-T controller is composed of polynomials { R_c, S_c, T_c }. At the identification stage, both current controllers C_{iq} and C_{id} should be selected previously.

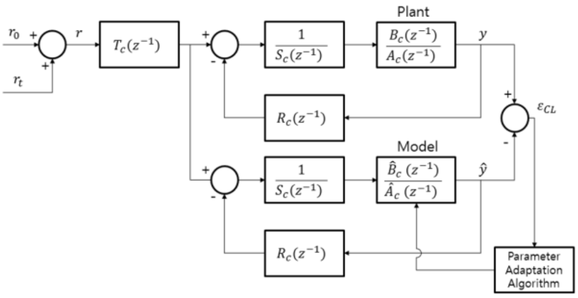

Figure 3: Closed-loop output error identification method

Letting $y(k) := i_q(k)$, the closed loop system is

$$y(k+1) = -\{A_c(q^{-1}) - 1\} y(k) + B_c(q^{-1}) u(k-d) + e(k) \tag{5}$$

$$S_c(q^{-1}) u(k) = T_c(q^{-1}) r(k) - R_c(q^{-1}) y(k) \tag{6}$$

The recursive adaptation algorithm of the CLOE method is briefly given as follows.

$$\begin{aligned} &\hat{\theta}(k+1) = \hat{\theta}(k) + F(k)\phi(k)\varepsilon_{CL}(k+1) \\ &F(k+1) = \lambda_1(k)F(k)^{-1} + \lambda_2(k)\phi(k)\phi(k)^T, \\ &\qquad (0 \le \lambda_1 \le 1; \quad 0 \le \lambda_2 \le 2) \\ &\varepsilon_{CL}(k+1) = y(k+1) - \hat{y}(k+1) \\ &\hat{y}(k+1) = \hat{\theta}(k)^T \phi(k) \end{aligned} \tag{7}$$

where

$$\hat{\theta}(k)^T := [\hat{a}_{c1}(k), \cdots, \hat{a}_{nac}(k), \hat{b}_{c1}(k), \cdots, \hat{b}_{nbc}(k)]$$

$$\phi(k)^T := [-\hat{y}(k), \cdots, \hat{y}(k-na+1), \hat{u}(k-1), \cdots, \hat{u}(k-nb+1)]$$

$$\hat{y}(k+1) := -\{\hat{A}_c(k, q^{-1}) - 1\}\hat{y}(k) + \hat{B}_c(k, q^{-1})\hat{u}(k)$$

$$S_c(q^{-1})\hat{u}(k) := T_c(q^{-1}) r(k) - R_c(q^{-1})\hat{y}(k) .$$

Depending on what operating conditions (either load disturbance or command speed and both) are considered, the value of r_o is changed. The identification is well converged if the amplitude of excitation input r_t is selected as about $\pm 10\,\%$ of the magnitude of r_o.

Similarly, a LTI model for the plant from v_d to i_d can be identified by superposing the test input r_t onto i_d^*. If the PMSM maintains a constant load and a constant low speed, both d- and q- axes models are similar each other.

Suppose that the equivalent plant G_c has been identified and the RST type digital controllers are designed from the models $\hat{G}_c(z^{-1})$. The design method will be presented in Section 3.

2.3 Experimental modeling of the plant in the speed control loop

As mentioned in previous subsection, once current tracking controllers are designed, they are implemented in the microprocessor. Then the current tracking control loop shown in Figure 4 can be described as an equivalent LTI plant, G_s. Thus the speed control system of PMSM is represented by a single loop feedback system, as shown in Figure 5.

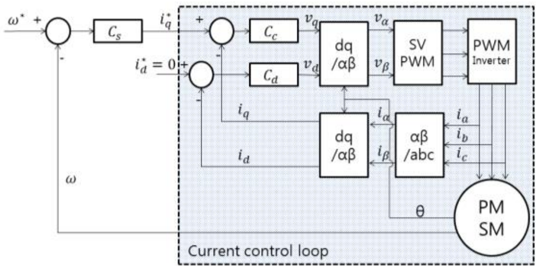

Figure 4: An equivalent plant in the speed control loop

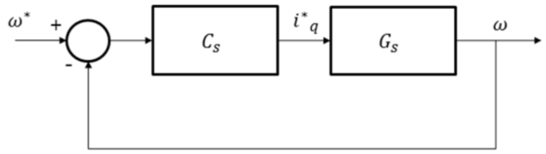

Figure 5: Block diagram of the speed control system in Figure 4.

It is seen that the problem of identifying the model G_s is the exactly same as that of identifying G_c presented in previous section.

An identified model of the discrete-time plant G_s is defined as

$$\hat{G}_s(z^{-1}) = \frac{\hat{B}_s(z^{-1})}{\hat{A}_s(z^{-1})}, \tag{8}$$

where

$\hat{A}_s(z^{-1}) = 1 + \hat{a}_{s1} z^{-1} + \cdots + \hat{a}_{nas} z^{-nas}$,

$\hat{B}_s(z^{-1}) = \hat{b}_{s1} z^{-1} + \cdots + \hat{b}_{nbs} z^{-nbs}$.

To apply the CLOE to the speed control loop, we need to select an arbitrary speed controller C_s provided that the closed loop is stable.

Comparing Figure 3 and Figure 5, if we let $y(k) := \omega(k)$, $r_o(k) := \omega^*(k)$, and use a RST type controller instead of C_s, the same CLOE algorithm as the method in Figure 3 can be applied for identifying $\hat{G}_s$. A PRBS test input r_t is superposed onto the reference speed $r_o(k) := \omega^*(k)$. Then this model is used for the analytic design of speed controller.

3 Direct design of digital controllers

It was shown that the vector control problem of the PMSM can be transformed by two independent single loop feedback control systems, as seen in Figure 2 and Figure 5. In this section, we concentrate on how the digital controller of each loop can be designed directly so that the resulting system satisfies the desired time response specifications: overshoot and settling time. The DCRA [6, 7] is used for this purpose.

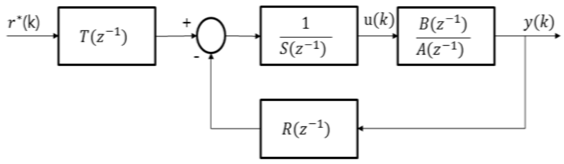

Figure 6: Feedback system with RST type controller

Consider the discrete-time feedback control system shown in Figure 6. An LTI plant and an RST type controller are described by

$$\frac{B(z^{-1})}{A(z^{-1})} = \frac{b_1 z^{-1} + b_2 z^{-2} + \cdots + b_{nb} z^{-nb}}{1 + a_1 z^{-1} + a_2 z^{-2} + \cdots + a_{na} z^{-na}}, \tag{9}$$

$$S(q^{-1})u(k) = T(q^{-1})r(k) - R(q^{-1})y(k), \tag{10}$$

where

$S(z^{-1}) = 1 + s_1 z^{-1} + \cdots + s_{ns} z^{-ns}$,

$R(z^{-1}) = r_0 + r_1 z^{-1} + \cdots + r_{nr} z^{-nr}$,

$T(z^{-1}) = t_0 + t_1 z^{-1} + \cdots + t_{nt} z^{-nt}$.

Since the plant model is assumed to be identified, the polynomials $A(z^{-1})$ and $B(z^{-1})$ are given.

The closed-loop system is given by

$$H(z^{-1}) = \frac{T(z^{-1})B(z^{-1})}{A(z^{-1})S(z^{-1}) + B(z^{-1})R(z^{-1})} = \frac{T(z^{-1})B(z^{-1})}{P(z^{-1})} \tag{11}$$

The characteristic polynomial $P(z^{-1})$ is

$$\begin{aligned} P(z^{-1}) &= A(z^{-1})S(z^{-1}) + B(z^{-1})R(z^{-1}) \\ &= p_0 + p_1 z^{-1} + \cdots + p_n z^{-n}, \end{aligned} \tag{12}$$

The DCRA is a model matching method. Let the desired reference model be $H^*(z^{-1}) = N^*(z^{-1}) \big/ P^*(z^{-1})$.

In the model matching approach, the controller {R,S,T} shall be determined such that $H(z^{-1}) \equiv H^*(z^{-1})$, that is, $T(z^{-1})B(z^{-1}) \equiv N^*(z^{-1})$ and $P(z^{-1}) \equiv P^*(z^{-1})$. Suppose that a reference model is given. The first step is to find the feedback term {R, S} of the controller polynomials by solving the following algebraic equation:

$$A(z^{-1})S(z^{-1}) + B(z^{-1})R(z^{-1}) = P^{*}(z^{-1}) \tag{13}$$

This identity equation has a unique solution if the following conditions hold:

$n_p \le n_A + n_B - 1,\;\; n_R \le n_A - 1,$ and $n_s \le n_B - 1.$

To achieve the zero steady state error to a step reference input, the overall system must be of *Type I*. From this condition, the feedforward term $T(z^{-1})$ can be obtained by

$$H(z^{-1})|_{z=1} = \frac{T(1)B(1)}{P(1)} = 1 \;\Rightarrow\; T(1) = \frac{P(1)}{B(1)}\;. \tag{14}$$

Now, the remaining problem is to find a reference model $H^{*}(z^{-1})$ that meets a prescribed transient response such as the maximum overshoot and settling time. Note that the problem of finding $H^{*}(z^{-1})$ is boiled down to the problem of finding $P^{*}(z^{-1})$ with a fixed numerator polynomial $N^{*}(z^{-1})$ if $T(z^{-1})$ of the degree zero is selected. The DCRA [6, 7] is a simple and very useful method to synthesize such a transfer function model.

For the design of vector controller of the PMSM, the DCRA is applied three times: to the current controllers { C_c , C_d } and to the speed controller, C_S . For example, the speed control system is shown in Figure 7.

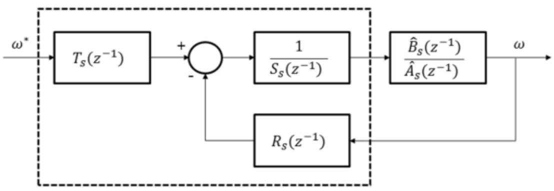

Figure 7: Speed control system with two parameter configuration and an identified plant model

4 Experimental results

Experiments on a 5 KW PMSM servo system have been carried out. Main parameters of the PMSM are listed in Table 1. A flywheel load of 220 Kg was mounted on the PMSM with a gear ratio of 5:1, as shown in Figure 8. The Digital controller including SVPWM, Clarke and Park's transformations is implemented by a DSP (TMS320F28335, TI) with a clock frequency of 150 MHz. The PMSM is driven by a three phase PWM inverter with an intelligent power module (PS21A7A, 600V, 75A, Mitsubishi). Figure 9 shows the configuration of the experimental setup.

Table 1: Parameters of the PMSM

Rated power	W	5000
Rated voltage	V	220
Rated speed	rpm	3000
Rated torque	N·m	15.9
Motor inertia	Kg· m^2	42.9x10^{-4}
Rated current	A	23.3
Encoder	point/r	2000

Figure 8: Experimental setup

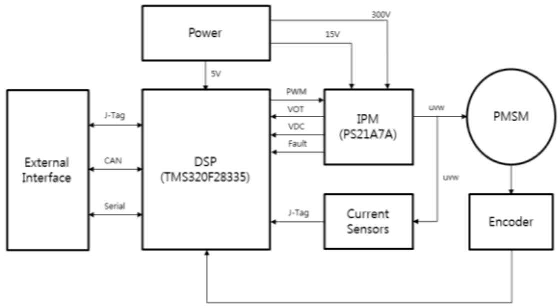

Figure 9: Configuration of the experimental setup

4.1 Modeling and controller design for the current loop

For modeling of the current loop, we consider 4 *q*-axis currents as different operating points: 3.5, 4, 5.5, and 7 A, respectively. The corresponding steady state motor speeds are about 500, 1000, 2000, and 3000 rpms. Among them, the modelling procedure at 5.5 A (about 2000 rpm) is represented here. The sampling time chosen for the current loop is 200 μs . In order to identify G_c using the recursive CLOE algorithm, we first select a temporary current controller C_c^o arbitrarily as follows: $S_c^o = 1 - z^{-1},\; R_c^o = 0.502 - 0.5z^{-1},\;\; T_c^o = 0.002$. A test input i_q^* and its time response i_q are shown in Figure 10. Through repeating identification processes on different orders of the model, it has been investigated that the following first-order model is well fitted to the experimental data.

$$\frac{\hat{B}_C(z^{-1})}{\hat{A}_C(z^{-1})} = \frac{\hat{b}_{c1} z^{-1}}{1+\hat{a}_{c1} z^{-1}} = \frac{0.05858 z^{-1}}{1-0.998 z^{-1}} \quad (15)$$

Figure 11 shows the parameter estimate of the model (15), which is obtained using the recursive CLOE algorithm (7). Time response of this estimated model is compared with the actual data in Figure 10.

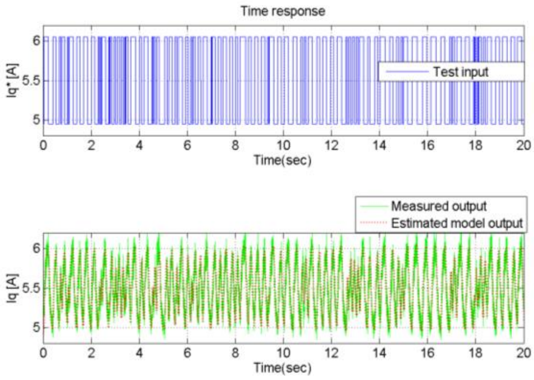

Figure 10: A PRBS test input applied to the q-axis reference current and its responses. Experimental data (green) and estimated model output (red).

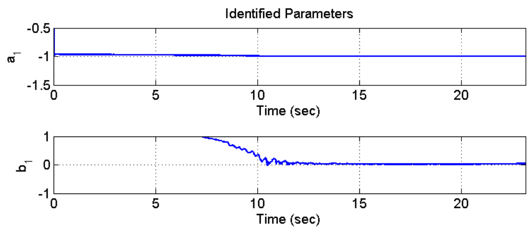

Figure 11: Estimated parameters of the q-axis current model at 2000 rpm using the CLOE algorithm.

Estimated parameters of the *q*-axis current model at four operating points are listed in Table 2.

Table 2: Estimated parameters of current loop model at 4 different operating points

	$i^*_{q0} = 3.5$ A (500rpm)	4A (1000rpm)	5.5A (2000rpm)	7A (3000rpm)
$\hat{a}_{c1}$	0.9963	0.9974	0.998	0.996
$\hat{b}_{c1}$	0.4726	0.05088	0.05858	0.09786

Similarly, models for the *d*-axis plant from v_d to i_d are identified. The result estimated at 2000 rpm is as follows:

$$\frac{\hat{B}_D(z^{-1})}{\hat{A}_D(z^{-1})} = \frac{\hat{b}_{d1} z^{-1}}{1+\hat{a}_{d1} z^{-1}} = \frac{0.04525 z^{-1}}{1-0.984 z^{-1}} \quad (16)$$

Now, we go to design a current tracking controller using DCRA. Suppose that the current tracking system is required to have the time response performances: non-overshoot and settling time of 50 *ms*. As explained in Section 3, the DCRA requires choosing the controller structure and a reference model satisfying the given specifications. From (15) and (16), a first order controller like PI type seems to be pertinent. In this paper, the PI type controller is selected for both current controllers. The order of characteristic polynomial can be determined. In DCRA, a reference model having the desired transient response is synthesized by selecting two design parameters: characteristic ratio (α_1), and a generalised time constant (τ). According to the rule [7], the reference characteristic polynomial $P^*_c(z^{-1})$ for the *q*-axis model in (15) is selected by

$$P^*_C(z^{-1}) = 1 - 1.967 z^{-1} + 0.9673 z^{-2}. \quad (17)$$

In other words, the step response of the reference model $H^*_C(z^{-1}) = {t_0 \hat{B}_C(z^{-1})}/{P^*_C(z^{-1})}$ has no overshoot and the exact settling time of 50 *ms*.

Replacing $P^*(z^{-1})$ of (13) by (17) and solving the algebraic equation (13), we have the following discrete-time PI controller C_c for the *q*-axis current loop:

$$S_C(z^{-1}) = 1 - z^{-1},$$
$$R_C(z^{-1}) = r_0 + r_1 z^{-1} = 0.5289 - 0.5231 z^{-1}, \quad (18)$$
$$T_C(z^{-1}) := t_0 = 0.0057.$$

Similar to the above controller, the *d*-axis PI controller C_d is obtained by

$$S_D(z^{-1}) = 1 - z^{-1},$$
$$R_D(z^{-1}) = 0.3765 - 0.369 z^{-1}, \quad (19)$$
$$T_D(z^{-1}) = 0.00749.$$

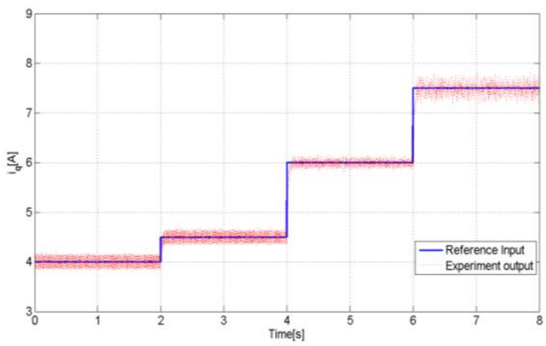

Figure 12: Response of a digital control in current loop for variable reference input.

The tracking performance of the resulting current controller $\{C_c, C_d\}$ is shown in Figure 12 for the case where the reference input i_q^* is changed stepwise as 4, 4.5, 6, 7.5 A at 2 sec intervals.

4.2 Modeling and controller design for the speed feedback loop

In the previous step, the current controllers have been obtained. They are implemented in the microprocessor. As explained in section 3.2, the current tracking control loop shown in Figure 4 can be described as an equivalent plant G_s in Figure 5. The sampling time chosen for the speed loop is $3\,ms$. In order to identify G_s using the recursive CLOE algorithm, we first select a temporary speed controller C_s^o arbitrarily as follows: $S_s^o = 1 - 0.5z^{-1}$, $R_s^o = 0.1$, $T_s^o = 0.1$. A test input ω^* and its time response ω are shown in Figure 13.

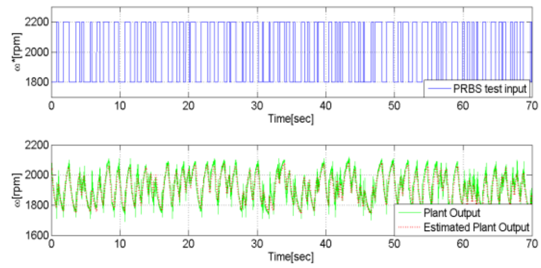

Figure 13: A PRBS test input and time response for speed loop. Estimated model response (red) and experiment output (green)

Applying the recursive CLOE algorithm to these data, we have the following second-order model for the speed loop.

$$\frac{\hat{B}_S(z^{-1})}{\hat{A}_S(z^{-1})} = \frac{0.1018z^{-1}}{1 - 0.4478z^{-1} - 0.552z^{-2}} \tag{20}$$

Time response of this estimated model is compared with the actual data in Figure 13. The parameter estimates obtained by the recursive CLOE method are shown in Figure 14.

We here consider that the speed control system is required to have the time response performances: non-overshoot and settling time of 3 sec to the step input. This limit is caused by the maximum value of the permissible currents on both hall sensors and the IPM module. We have observed through simulations that it is difficult for this experimental setup with a PID type controller to have good performance over the full range of operating speed. So, a second-order RST controller including an integrator has been selected.

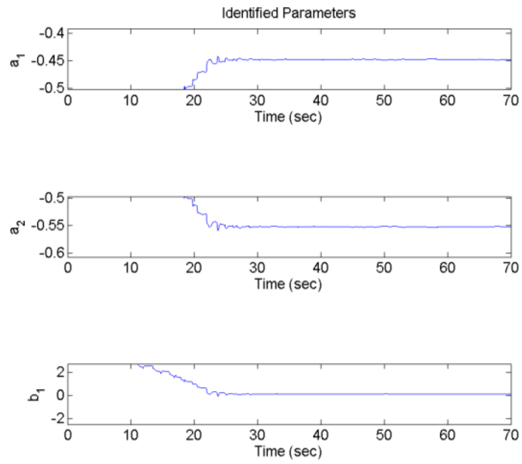

Figure 14: Parameter estimates of speed model

According to the rule [7] again, the reference characteristic polynomial $P_S^*(z^{-1})$ for the speed model in (20) is selected by

$$P_s^*(z^{-1}) = 1 - 1.98585z^{-1} + 0.68155z^{-2} + 0.62267z^{-3} - 0.31829z^{-4} \tag{21}$$

Replacing $P^*(z^{-1})$ of (13) by (21) and solving the algebraic equation (13), we have the following RST type speed controller C_S.

$$\begin{aligned} S_s(z^{-1}) &= (1 - z^{-1})(1 - 0.5767z^{-1}), \\ R_s(z^{-1}) &= 0.38029 - 0.4851z^{-1} + 0.10564z^{-2}, \\ T_s(z^{-1}) &:= t_0 = 0.00076. \end{aligned} \tag{22}$$

This speed controller is implemented in the DSP. Now, the overall system including both current and speed controllers is demonstrated experimentally. When the reference speed is changed stepwise, the time response of speed control satisfies the design requirements very well, as shown in Figure 15.

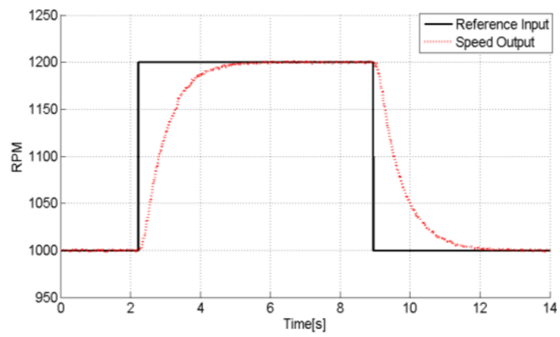

Figure 15: Result of speed control

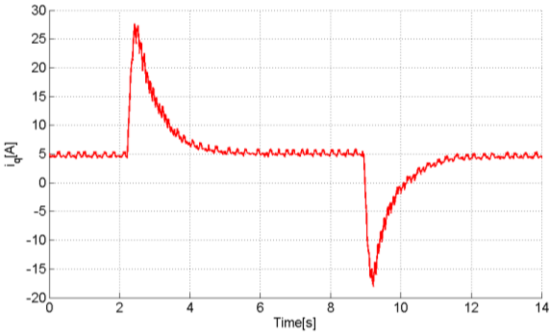

Figure 16: Result of current control

5 Conclusions

Digital control system of the PMSM based on a mathematical model requires the identification of physical parameters, which indicate electrical characteristics and mechanical motion. However, it is not easy to know some of these parameters (e.g., inertia moment and friction coefficient) exactly. Furthermore, they are dependent upon load and operating conditions.

In this paper, the input-output modeling technique based on the closed loop identification has been considered to estimate the equivalent discrete-time models. Modeling and controller design are performed for the current control and speed control loops separately. After estimating the current loop model, a discrete-time current controller based on the model is designed so that the tracking performance satisfies the desired transient responses; maximum overshoot and settling time. Next step is to identify the equivalent speed model including the current controller. Based on this estimated model, a speed controller is directly designed using the DCRA under the given time response requirements.

The proposed approach has been demonstrated by experiments. The experimental setup consists of a surface mounted PMSM (5 KW, 3ϕ 220V, 8 poles, Higen) equipped with a flywheel load of 220Kg, a three phase PWM inverter with an IPM (PS21A7A, 600V, 75A, Mitsubishi), and a digital controller using DSP (TMS320F28335). Modeling for the current loop has been performed at four operating points and results in good outcomes. It was shown that all the experimental responses coincide closely with those of reference model.

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Authors

Taesu Kang received the B.S degree in information and communication eng. and M.S. degree in electronics eng. from the Chungbuk National University, Cheongju Korea, in 2013 and 2015, respectively. His research interests include digital control system design for Motor drive and obstacle detection for autonomous vehicle.

Min-Seok Kim received the B.S degree and M.S. degree in electronics eng. from the Chungbuk National University, Cheongju Korea, in 2005 and 2014, respectively. He was a senior researcher at the Samsung Elec. Co. Ltd. (2005–2009) and RIUBIT, (2009-2012). He works currently at the Central Research Center, Korea Hydro & Nuclear Power Co. Ltd.

Kiyong Lee received the B.S degree in electronics eng. from the Chungbuk National University, Cheongju Korea, in 2014, where he is currently pursuing the M.S. degree.

Sa Young Lee received the M.S. and Ph.D. degrees in electrical eng. from the Myongji University, Seoul Korea, in 1982 and 1992, respectively. He is currently a professor with Department of Elec. Eng., Myongji College. His research interests are concentrated on theory and applications of power electronics eng.

Young Chol Kim received the B.S. from Korea University Seoul Korea in 1980, M.S. and Ph.D. in electrical eng. from the Seoul National Univ., Seoul Korea, in 1982 and 1987, respectively. He was a visiting professor at the Texas A&M Univ. (1991) and Vanderbilt Univ. (2001).
He is a professor at School of ECE, Chungbuk National Univ. Cheongju, Korea since 1988.