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Full Length Research Paper

Permanent magnet flux estimation method of vector control SPMSM using adaptive identification

Abdoulaye M'bemba Camara*, Yosuke Sakai and Hidehiko Sugimoto

Department of Electrical and Electronics Engineering, University of Fukui, Fukui 910-8507, Japan.

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The variation of permanent magnet flux deteriorates the performance of a torque controlled (TC) system. Without the torque sensor, the magnet flux information is indispensable for controlling the torque of the surface permanent magnet synchronous motor (SPMSM) with vector control. The magnet flux depends on variations of temperature inside of the motor. With the increase of stator winding temperature, the magnet temperature increases and the magnet flux decreases. Therefore, the magnet flux is not treated constantly and thus becomes a big issue to TC. So, the instantaneous value of the magnet flux is needed in any way. This paper proposes the estimation method of the magnet flux based on the adaptive identification used in the SPMSM with the vector control. Even at low speed, the influence of the stator resistance variation is not received easily because the proposed method has the stator resistance estimation function. The effectiveness of the proposed method is demonstrated by experiments.

Key words: SPMSM, magnet flux, vector control, stator resistance estimation, adaptive identification, mathematical model.

INTRODUCTION

The surface permanent magnet synchronous motor (SPMSM) drive system has been widely used for high efficiency application. In recent years, torque controlled (TC) systems have become one of the favored control schemes for induction machine. In this case, controlling amplitude of the armature currents is required. The same principle has been applied in interior permanent magnet synchronous motors (Rahman et al., 2003; Tang et al., 2003).

Some AC servo motors used for injection molding machines (Yoshiharu et al., 1999) require torque control with less than 1% error. Such motors are SPMSMs in field system that has Neodym magnet, an high energy product, but torque feedback system is needed to detect torque with torque sensor. With the increase of the stator winding temperature, the magnet temperature increases and then the magnet flux decreases; and as a result, the torque decreases proportionally.

In Takeshit et al. (1997) and Beccera et al. (1991), an observer of the stator resistance variation is added to decrease the negative effects of the stator resistance variation. However, the stator resistance of the SPMSM is affected by many factors of the environment. Thus, it is difficult to estimate the resistance variation accurately on line, and the magnet flux of the motor cannot be estimated accurately with temperature increase. Some researchers have developed observers using the motor model (Han et al., 2000; Fukumoto et al., 2007; Hayashi et al., 2005). A suitable design of the observers produces a high level of insensitivity to parameter variations, but the observers are sensitive to noise measurement; the operation load seems to be large and their stability analysis is difficult because of many expressions.

During the low speed operation, the contact time between machine and plastic, injection molding machines, as an example, is important. That is why it is necessary to estimate the magnet flux during this period because much current would flow and cause a very important variation of temperature in the armature winding of SPMSM. Since the torque is necessary in the injection molding machine during the operation, it is constant even

*Corresponding author. E-mail: aziz71@u-fukui.ac.jp. Tel: +81-776-23-0550. Fax: +81-776-27-8749.

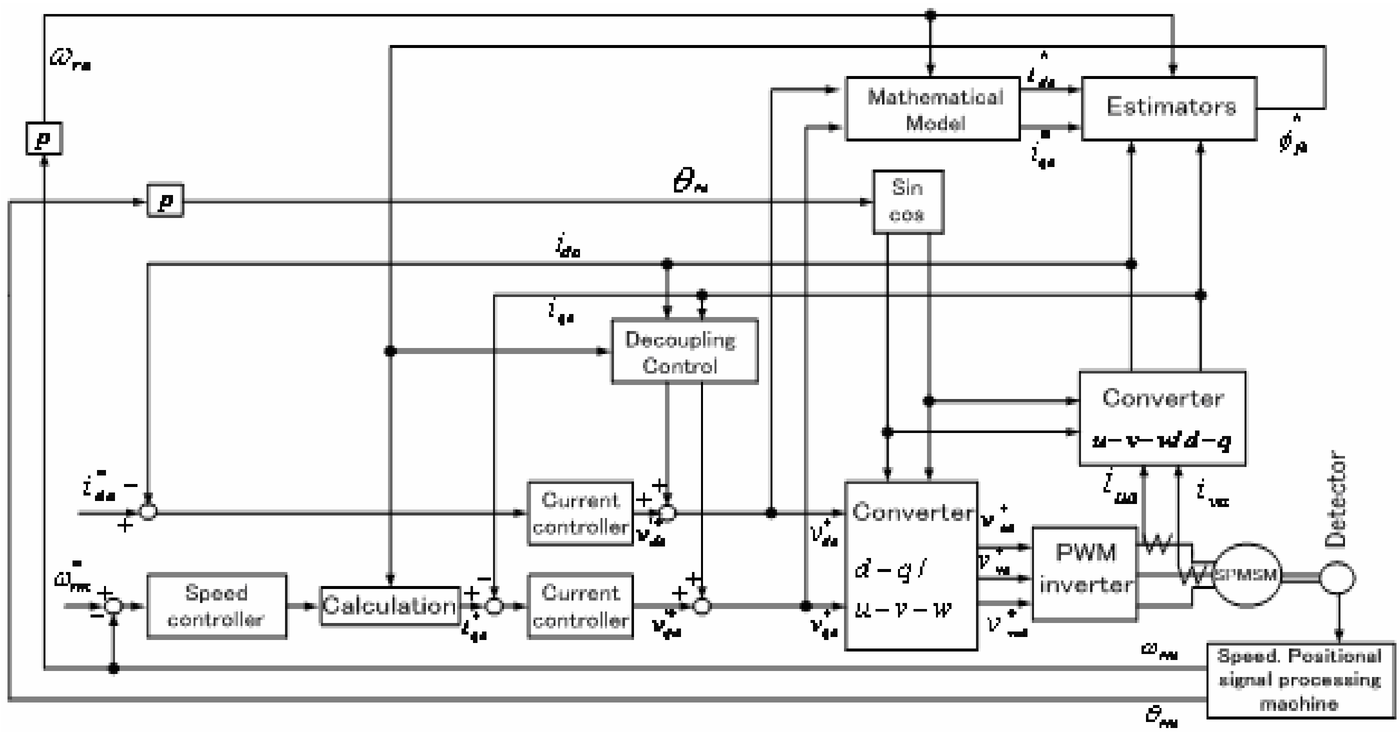

Figure 1. Block diagram of the control scheme.

at low speed; therefore, the magnet flux estimation becomes necessary for torque control systems.

The research about the temperature sensor is done, but the research about high accuracy of torque control without torque sensor is not active.

We propose a new torque control system without torque sensor controlling armature current with error less than 1% by using adaptive identification of magnet flux with temperature variation. The salient features of the new method are as follows:

(1) The magnet flux and the armature winding resistance are estimated by integrating the magnet flux estimation error with the armature winding resistance estimation error, leading to the derivation of the linearization error state equation of a real machine and using of a mathematical model.
(2) The estimation of the magnet flux is not influenced easily from estimation of the armature winding resistance.
(3) The design of the magnet flux estimator is simple; actually the design of bandwidth of closed loop transfer function between value and estimator of magnet flux is simple equation first–degree.
(4) For adjustment of moving average at low speed to estimate magnet flux.

In this thesis, the linearization error margin of state equation from state equation of a real machine and the mathematical model and the estimator is composed by using it. The stability is examined by bode diagram, and the utility is confirmed by the experiment.

MATERIALS AND METHODS

Mathematical model and linearization error equation of state

Composition of the proposed system

Figure 1 shows the composition of the proposed system. The instruction value is shown as * in Figure 1. The voltages, v_{da} , v_{qa} are given as real SPMSM; and mathematical model as an input, from which the mathematical model estimates, $\hat{\phi}_{fa}$ $\hat{R}_a$; the quantities, $\Delta\phi_{fa}, \Delta R_a$ are causes of the current estimation error. And these may become 0 by using the estimation of the output current.

Mathematical model

The state equation of SPMSM in (d, q) coordinates is shown in Eq. (1). Moreover, Eq. (2) shows the mathematical model in Figure 1 by which the estimated current is described; $\hat{i}_{da}$ and $\hat{i}_{qa}$ are estimated by using $i_{da}, i_{qa}, v_{da}, v_{qa}, \hat{\phi}_{fa}, R_a$

$$\begin{bmatrix} L_a & 0 \\ 0 & L_a \end{bmatrix}\frac{d}{dt}\begin{bmatrix} i_{da} \\ i_{qa} \end{bmatrix} = \begin{bmatrix} -R_a & \omega L_a \\ -\omega L_a & -R_a \end{bmatrix}\begin{bmatrix} i_{da} \\ i_{qa} \end{bmatrix} + \begin{bmatrix} v_{da} \\ v_{qa} \end{bmatrix} - \omega_{re}\phi_{fa}\begin{bmatrix} 0 \\ 1 \end{bmatrix} \quad \text{.......... (1)}$$

$$\begin{bmatrix} L_d & 0 \\ 0 & L_d \end{bmatrix} \frac{d}{dt} \begin{bmatrix} \hat{i}_{da} \\ \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} -\hat{R}_a & \omega L_d \\ -\omega L_d & -\hat{R}_a \end{bmatrix} \begin{bmatrix} \hat{i}_{da} \\ \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} v_{da} \\ v_{qa} \end{bmatrix} - \omega_{re} \hat{\phi}_{fa} \begin{bmatrix} 0 \\ 1 \end{bmatrix} + \begin{bmatrix} g_{11} & g_{12} \\ g_{21} & g_{22} \end{bmatrix} \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} \quad (2)$$

R_a : stator winding resistance L_a :d,q axes inductances, ϕ_{fa} permanent magnet flux.

v_{da}, v_{qa} :d,q axes voltages, i_{da}, i_{qa} :d,q axes currents , ω_{re} :synchronous angular frequency.

In this work, the current of the armature and the parameter with the sign, ^ show "estimation" and the estimated parameter

are ϕ_{fa} , R_a . Moreover, $g_{11}, g_{12}, g_{21}, g_{22}$ are mathematical model gains.

Linearization of estimation error of state equation

The following state equation for estimation error is obtained by multiplying matrix inverse of inductance matrix to each side of the Eqs (1) and (2):

$$\frac{d}{dt} \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} B_{11} & B_{12} \\ B_{21} & B_{22} \end{bmatrix} \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix} = Ae_{ia} + Bu \quad (3)$$

Where:

$$A_{11} = -\left(\hat{R}_a + g_{11}\right) / L_a, A_{12} = (\omega_{re} L_a - g_{12}) / L_a, A_{21} = (-\omega_{re} L_a + g_{21}) / L_a$$

$$A_{22} = -\left(\hat{R}_a + g_{22}\right) / L_a, B_{11} = 0, B_{12} = -i_{da} / L_a, B_{21} = -\omega_{re} / L_a, B_{22} = -i_{qa} / L_a$$

$$e_{ia} = \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix}, u = \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix}$$

The above linear equation consists of the armature current, the armature voltage, the turning angle speed, and the armature winding resistance. The linearization makes the first separate equilibrium point (hereafter, it is shown that subscript 0 with lower right is an equilibrium point) and changes mathematical model gain when approximating. Each equilibrium point of the Eqs (1) and (2) is assumed to take the same value:

$$s \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} A_{110} & A_{120} \\ A_{210} & A_{220} \end{bmatrix} \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} B_{110} & B_{120} \\ B_{210} & B_{220} \end{bmatrix} \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix} = A_0 e_{ia} + B_0 u \quad (4)$$

The mathematical model gain of Eq. (5) is chosen.

$$g_{11} = g_{22} = g_a, g_{12} = \omega_{re} L_a, g_{21} = -\omega_{re} L_a \quad (5)$$

The characteristic equation of the mathematical model is shown in Eq. (6):

$$(sI - A_0) = \begin{bmatrix} s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} & 0 \\ 0 & s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \end{bmatrix} \quad (6)$$

The mathematical model gain, g_a , which Eq (6) stabilizes, is chosen. Moreover, the linearization estimation error becomes as follows:

$$\begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = (sI - A_0)^{-1} B_0 \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix}$$

$$= \begin{bmatrix} 0 & \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{i_{da0}}{L_a} \right) \\ \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{\omega_{re0}}{L_a} \right) & \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{i_{qa0}}{L_a} \right) \end{bmatrix} \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix}$$

$$= P_0(s) \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix} \quad (7)$$

From where the transmission function matrix $P_0(s)$ is:

$$P_0(s) = \begin{bmatrix} 0 & \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{i_{da0}}{L_a} \right) \\ \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{\omega_{re0}}{L_a} \right) & \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{i_{qa0}}{L_a} \right) \end{bmatrix} \quad (8)$$

The inversion of $P_0(s)$ is:

$$P_0^{-1}(s) = \frac{\tilde{P}_0(s)}{\det(P_0)}$$

$$P_0^{-1}(s) = \begin{bmatrix} \left(s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}\right)\left(\frac{L_a i_{qa0}}{\omega_{re0} i_{da0}}\right) & \left(s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}\right)\left(-\frac{L_a}{\omega_{re0}}\right) \\ \left(s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}\right)\left(-\frac{L_a}{i_{da0}}\right) & 0 \end{bmatrix} \quad (9)$$

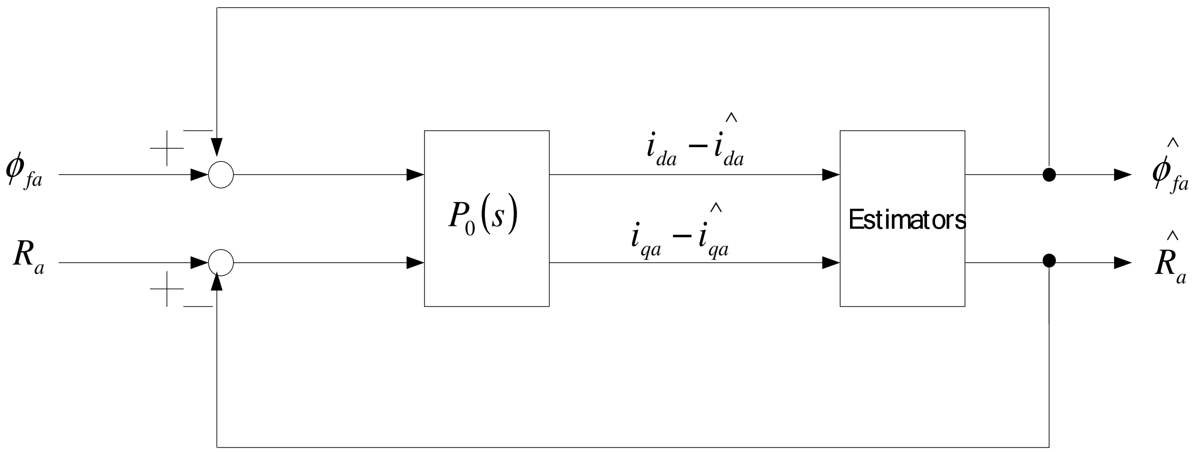

Figure 2. Construction of estimation system.

Influence of parameter errors

Controlling torque of SPMSM is usually done with torque sensor. However, we can reduce the cost by making it sensorless. The estimation of the magnet flux, which depends on temperature, is therefore necessary and indispensable. The magnet type of the tested motor is Nd-Fe-B and the temperature coefficient is about $0.11\ \% / {}^{0}C$. When the temperature of the motor changes to ±75º*C*, the variation of the permanent magnet flux is within $\mp$ 9% and the armature winding resistance changes to about ± 24%. In order to be related, occasionally the magnet flux is required in order to make a direct relation between the temperature and the magnet flux. With this proposition, the armature winding resistance and the magnet flux, which depend on temperature, are estimated; it is something which assures improvement of torque efficiency by the fact that the magnet flux estimator is used for control.

COMPOSITION OF ESTIMATOR

Construction of estimation system

The construction of an estimation system is shown in Figure 2 and it showed the l

$$\begin{bmatrix} \hat{\phi}_f \\ \hat{R}_a \end{bmatrix} = P_0^{-1}(s) \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} \quad \text{....... ....(10)}$$

And the estimator is:

$$\begin{bmatrix} \hat{\phi}_f \\ \hat{R}_a \end{bmatrix} = \begin{bmatrix} \left(1+\frac{1}{s}\frac{\hat{R}_a+g_a}{L_a}\right)\left(\frac{K_{p1}L_a i_{qa}}{\omega_{re} i_{da}}\right) & \left(1+\frac{1}{s}\frac{\hat{R}_a+g_a}{L_a}\right)\left(-\frac{K_{p2}L_a}{\omega_{re}}\right) \\ \left(1+\frac{1}{s}\frac{\hat{R}_a+g_a}{L_a}\right)\left(-\frac{K_{r1}L_a}{i_{da}}\right) & 0 \end{bmatrix} \begin{bmatrix} i_{da}-\hat{i}_{da} \\ i_{qa}-\hat{i}_{qa} \end{bmatrix} \quad \text{...... (11)}$$

Here, ω_{re} and i_{da} exist in the denominator when paying attention to the element of the first column of the first row. This makes i_{da} an important parameter, because if it is zero we will not able to estimate the magnet flux and the stator winding resistance. The same consequences will happen when the speed becomes zero, that is $\omega_{re} = 0$. However, from the viewpoint of speed resolution, estimation is not possible even at speed near to zero, that is, below the nominal speed.

In both Eqs (7) and (9), we get:

$$\begin{bmatrix} \hat{\phi}_{fa} \\ \hat{R}_a \end{bmatrix} = \begin{bmatrix} \frac{K_{p20}}{s} & \frac{(-K_{p10}+K_{p20})i_{qa0}}{\omega_{re0}s} \\ 0 & \frac{K_{r10}}{s} \end{bmatrix} \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix} \quad \text{............. ..(12)}$$

Where K_{p10}, K_{p20} and K_{r10} represent each integrator gain in Eq. (12). The second element of the first row becomes zero with $K_{p10} = K_{p20}$. We thus have the following formulation (Figure 2):

$$\begin{bmatrix} \hat{\phi}_{fa} \\ \hat{R}_a \end{bmatrix} = \begin{bmatrix} \frac{K_{p20}}{s} & 0 \\ 0 & \frac{K_{r10}}{s} \end{bmatrix} \begin{bmatrix} \Delta\phi_{fa} \\ \Delta R_a \end{bmatrix} \quad \text{.............................. (13)}$$

$$\begin{bmatrix} \hat{\phi}_{fa} \\ \hat{R}_a \end{bmatrix} = \begin{bmatrix} \frac{K_{p20}}{s+K_{p20}} & 0 \\ 0 & \frac{K_{r10}}{s+K_{r10}} \end{bmatrix} \begin{bmatrix} \phi_{fa} \\ R_a \end{bmatrix} \quad \text{........................ (14)}$$

These two equations are preferable when they are stable and are used to adjust the integrator gain to the satisfaction derived from value response. This is because the Eq. (14) is derived from the Eq. (13). The advantage of this method is robustness of estimation, because the influence of the armature winding resistance

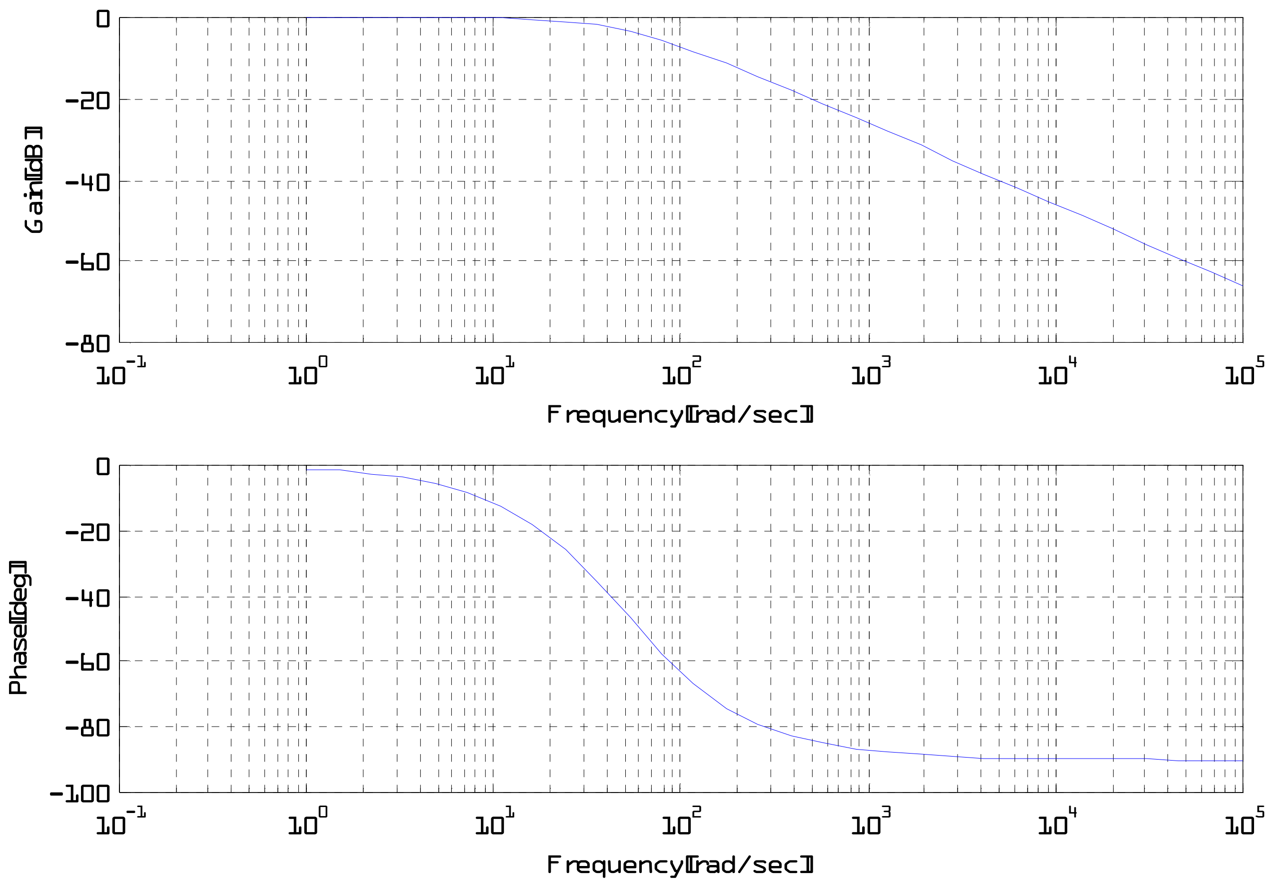

Figure 3. Bode diagram of $\hat{\phi}_{fa} / \phi_{fa}$.

estimation error ΔR_a is not easily received in the magnet flux estimation. The above two equations for the open-loop and the closed-loop transfer functions in the Eq. (13), (14) respectively work only at high-speed from where we can estimate the magnet flux. However, Eq. (1) can be used at speed near to zero because we cannot estimate the magnet flux and use it. First of all, we need to find when the speeds rotate faster than resolution. At that point, the open-loop transfer function expressed by Eq. (13) can estimate the magnet flux

Figure 3 shows the characteristic of the stability and it is realized at $K_{p20} = 50$. When a current condition is applied, it is arranged like Eqs. (13) and (14). The gain and phase differences between the input and output of the estimation system are a simple pole. The following observations can be made from Figure 3.

-That for a simple real pole, the piecewise linear asymptotic Bode plot for magnitude is at 0dB until the break frequency; and then drops at 20dB per decade (the slope is -20dB/decade).

-The phase plot is at 0° until one tenth the break frequency, and then drops linearly to -90º at ten times the break frequency.

Improvement of the speed resolution

The problem of low-speed rotating is recorded as follows. The output $\hat{\phi}_{fa}$ of the estimator might become unstable when it is more low-speed than Eq. (15). The Eq. (15) becomes low-speed when the torque is output as for the injection molding machine. Therefore, it is necessary to estimate the magnet flux until the speed is near to zero. Under this condition, there is a method of improving speed resolution with increase of the number of moving average samples. The calculated example of the improvement of the speed resolution $\Delta \omega_{rm}$ (mechanical speed) is recorded as follows.

$$\Delta\omega_{rm} = \frac{2\pi}{n_p \bullet t_c \bullet N} [\text{rad/sec}] \quad \text{..........................} (15)$$

Here n_p [pulse/rev] : the encoder pulse number t_c [μsec] : operational period, N number of moving average samples. $n_p = 4000$ [pulse/rev], $t_c = 204.8$[μsec] , $N = 64$.

Eq. (15) shows the improvement of the speed resolution with an increase in the number of samples of moving averages. That is necessary because even at speed near to zero, we can detect the speed, and the flux estimation becomes possible. The reason to use Eq. (15) is to obtain the operation accuracy even when there is a sudden change in speed. At this point, we need to know where the estimation should stop. This is because the flux estimation

Table 1. Rating of tested motor.

Rated power		**1.5kW**
Rated current		8.6 A
Rated speed		2000rpm
Number of pole-pairs	p	3
Torque constant	ϕ_{fa}	0.1946Wb
Armature winding resistance	R_a	0.5157 Ω
Armature winding self-inductance	L_a	2.452mH
Moment of inertia	J	0.00525 $kg \cdot m^2$

cannot be done at speed slower than the resolution speed.

Simulation and experimental conditions

A test system was composed of a Digital Signal Processor DSP (TMS320C31-5kHz) control system (Texas instruments), a 3-phase PWM inverter and a 1.5 kW SPMSM. The operation cycle is 200 μs ; the career frequency of the PWM inverter that drove the evaluation machine was assumed to be 5 KHz. The load machine used is the induction motor of 1.5kW and moment of inertia is 0.009 $kg \cdot m^2$. The detected current and voltage are fed to the input of DSP and then the voltage order was calculated. The voltage instruction from DSP is converted into the PWM signal, and then the short-circuit prevention time is added by FPGA which generates output to the circuit of the drive at the gate. Signal carrier (triangular wave) cycle was assumed to be 204.8 μs using the triangular wave comparison method for the generation of the PWM signal. Hall CT (HAS-50S: LEM) was used for the current detector.
The voltage proportional to the current from hall CT is output, and the voltage signal is converted into the digital signal with 16 bit A/D converter (AD976:AnalogDevices). DC power voltage E_{DC} of the inverter is detected with 12 bit A/D converter (AD7864: Analog Devices) connected through the partial pressure machine.
Voltage type PWM inverter is composed of the power-module and the circuit of the drive at the gate. IGBT-IPM (6MBP30RH060: Fuji Electric Co., Ltd.) was used for the power-module. The direct current voltage power supply of the inverter has vector control of faction 2.2kW of the three-phase circuit 200V type inverter (FRN2.2VG7S-2: Fuji Electric Co., Ltd.) that controls the torque; and DC linked the load machines. The pulse number output from the encoder used this time is 1000 pulses per rotation.
As for the voltage detection error margin $\pm 1/(2^{13}) = \pm 1/8192$, the delay of the voltage feed back loop becomes $300\mu s$, which is 1.5 times at sampling period $204.8\mu s$.

Rating of tested motor used in this experiment and simulation are in Table 1.

As for injection molding machine, the torque is particularly important especially at low speed and the estimation of the magnet flux at low speed becomes necessary.

With the increase of the stator winding temperature, the stator winding resistance increases and the magnet flux decreases as shown in Figure 4. The magnet flux estimation with robustness was confirmed by the simulation based on the condition described in Table 1. This is because stator winding resistance is not estimated. The simulation has been made through the language of technical computing (MATLAB) that omitted the PWM inverter.

The estimation of magnet flux in the simutation started at 5 s. Also, during 10 s, the temperature in the real machine would not vary as shown in the simulation.

The simulation results in Figure 4 show that when the temperature increases, the armature winding resistance increases and the magnet flux decreases.

RESULTS

Figures 5, 6 showed the simulation results of estimated magnet flux with 25% load at 20 and 200 rpm when the temperature as well as the stator winding resistance does not change. The initial value is fixed at 0.295 Wb and the estimation started at 0.8 s.

The result of the flux estimation at 20 rpm with no-load by the experiment is recorded in Figure 7. The result is similar to that of the simulation- when the temperature and stator resistance do not change.

The estimation of flux at 100% load is easy to realize at low speed during the experiment but it is more difficult to realize the result at 25% load. Therefore, it showed the result of experiment at 25% load during 20 and 200 rpm (Figures 8 and 9).

This experiment shows that, the estimation of magnet flux was effective because the voltage sensor is used in the experimental operation and hence the estimation error is 0.3%, less than 0.5%. Also, these results showed the performance of the experiment compared to the results of simulation is similar (having the same condition).

DISCUSSION

The magnet flux information is important for controlling the torque of the SPMSM with vector control. We propose in this work a new method for estimating the magnet flux of the SPMSM with vector control. The proposed method showed good estimated performance by designing and

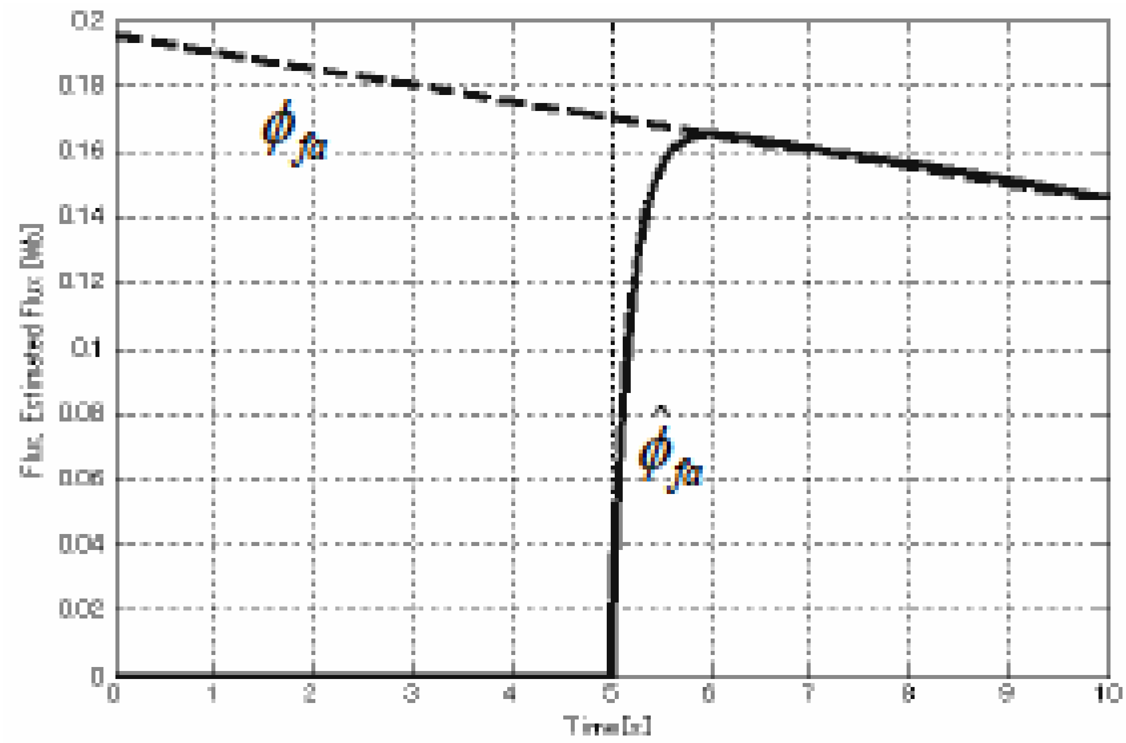

Figure 4. Simulation results for the flux and estimated flux.

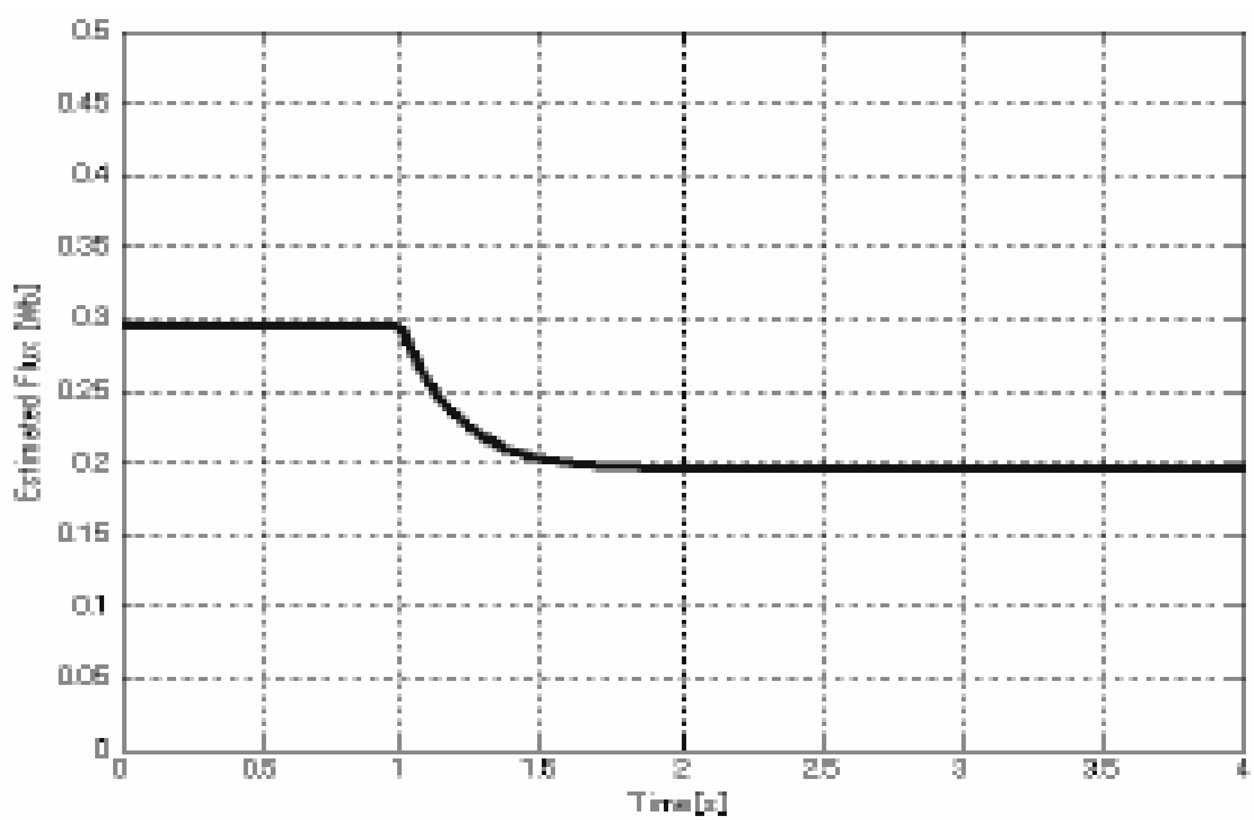

Figure 5. Simulation result for estimated magnet flux (25% load, 20 rpm).

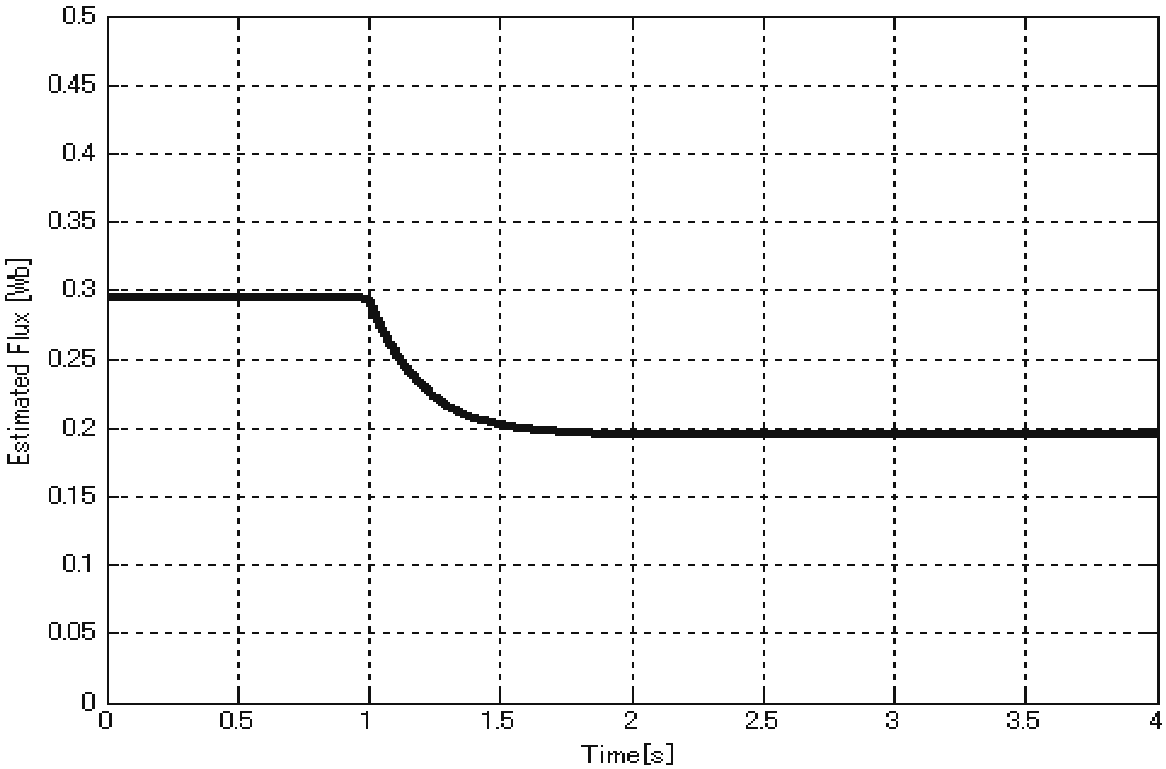

Figure 6. Simulation result for estimated for magnet flux (25% load, 200 rpm).

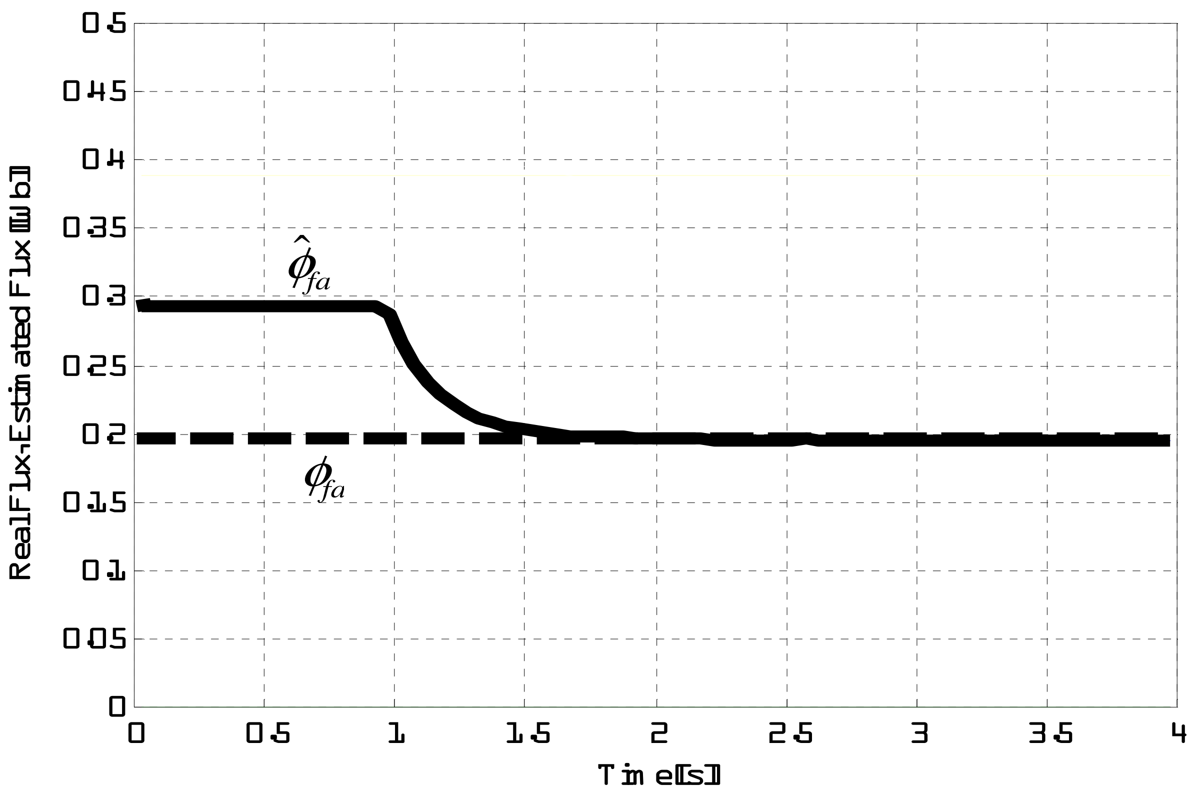

Figure 7. Experimental results for estimated of magnet flux (no-load, 20 rpm).

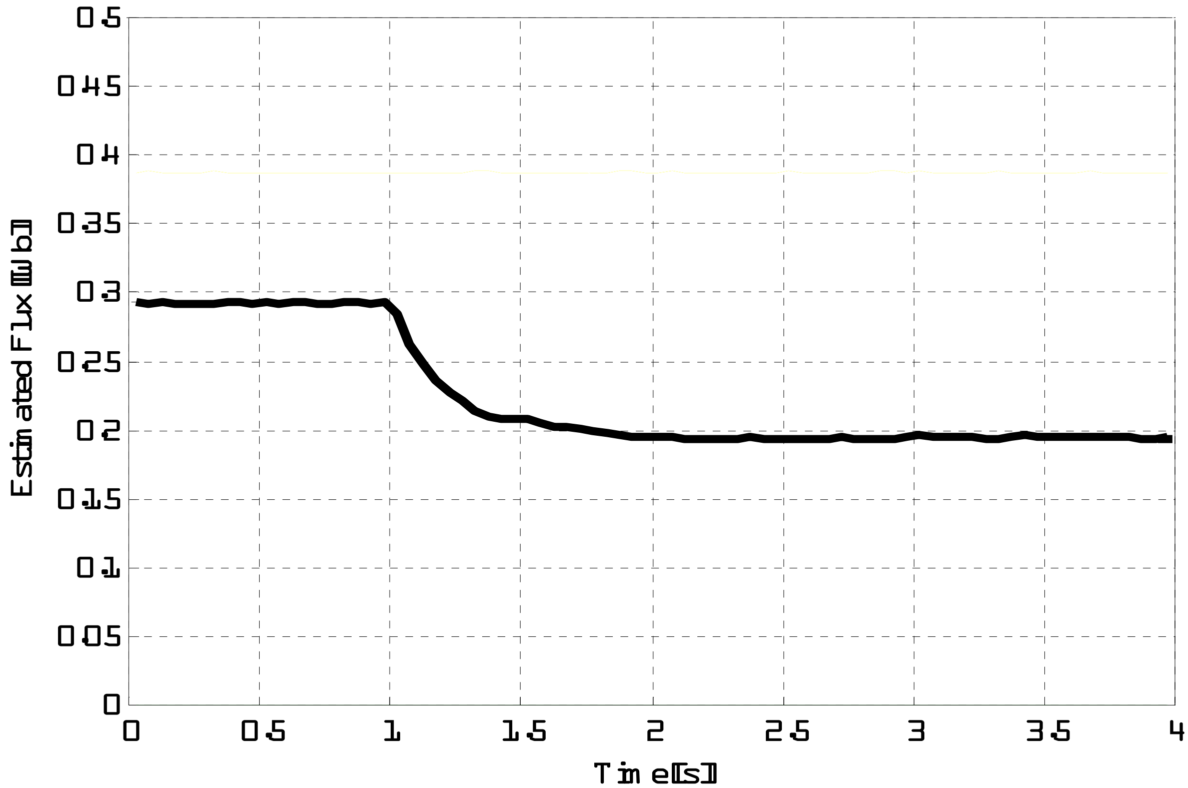

Figure 8. Experimental results for estimated of magnet flux (25% load, 20 rpm).

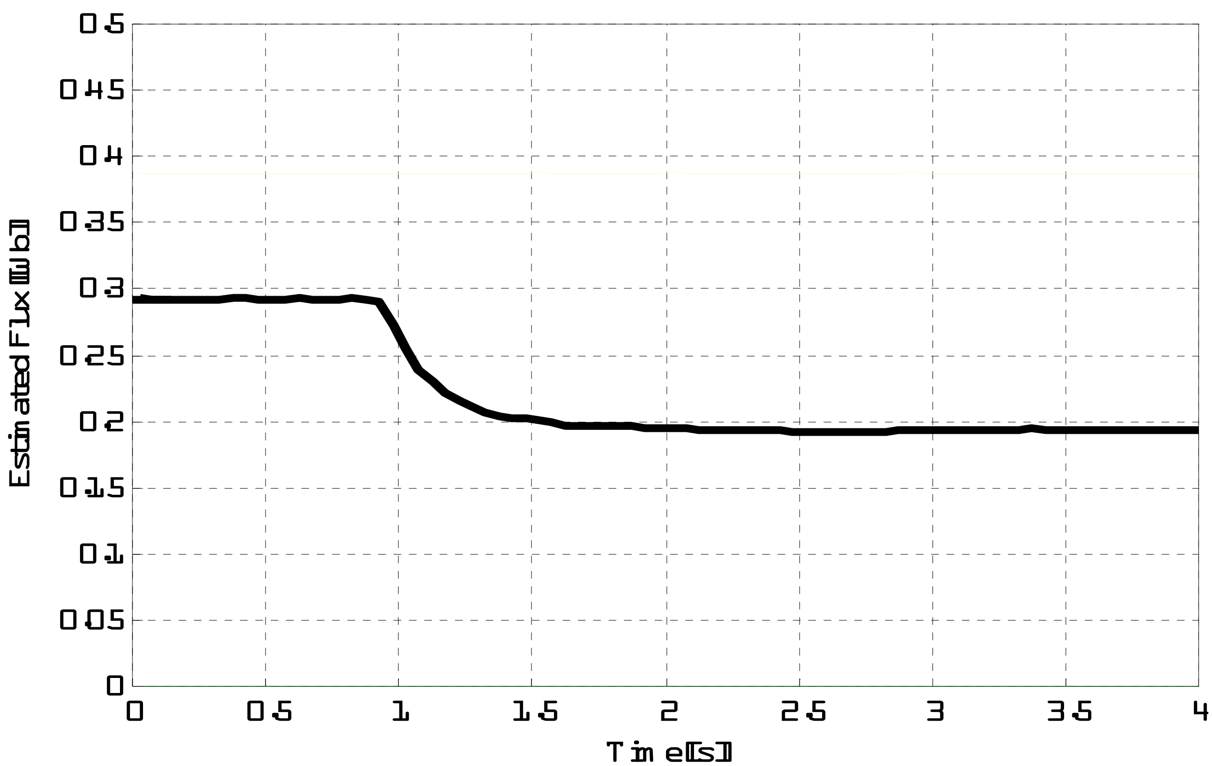

Figure 9. Experimental results for estimated of magnet flux (25% load, 200 rpm).

simulating the estimator of the magnet flux from linearization error state equation and also the robustness of the estimated magnet flux is demonstrated by noninterference of the stator winding resistance variation caused by the increase of the temperature in SPMSM. Currently, the condition of the experiment and starting flux estimation were not the same value. However, in future, we like to use the same value for both cases.

Conclusion

Since the torque is necessary in the injection molding machine during operation, it is constant even at low speed; and so the magnet flux estimation becomes necessary for torque control systems without torque sensor during the variation of stator winding resistance. Therefore, this paper presents good estimated performance by designing and simulating the estimator of the magnet flux with vector control using adaptive identification from linearization error state equation and also presents the robustness demonstrated by noninterference of the stator winding resistance variation caused by the increase of the temperature in SPMSM.

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