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A Study on Accurate Torque Control of Surface Permanent Magnet Synchronous Motor without Torque Sensor

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ABSTRACT

The variation of permanent magnet flux deteriorates the performance of 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 armature winding temperature, the magnet temperature increases and then the magnet flux decreases. Through that variation, the magnet flux is not treated constantly and then the magnet flux information becomes necessary to keep the pressure constant during the operation of machines as the injection molding machine and hence, the magnet flux becomes a big issue of TC. So, the instantaneous value of the magnet flux is needed in any way. Therefore, it is important to develop a fine force-control system. Generally, in force-control systems, the force information from the environment is detected by a force sensor. However, control systems using force sensors present problems related to signal noise, sensor cost, narrow bandwidth, and other factors. To overcome these problems, this thesis proposes the estimation method of the magnet flux of SPMSM based on the adaptive identification 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 armature winding resistance estimation function. The effectiveness of the proposed method is demonstrated by both of simulations and experiments.
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DEDICATION

I would like to dedicate this dissertation to my lovely wife Fatoumata Sampil for supporting me during all my studies here in Japan. At the bottom of my heart, thanks my love for all.
CHAPTER 1

GENERAL INTRODUCTION

1.1 ELECTRICAL MOTOR CONTROL

Electrical motors operate on the principle that two magnetic fields within certain prescribed areas react upon each other. All electric motors use electromagnetic fields to create torque. For many motion engineers, motor selection plays a central part in getting good device performance. Knowing which motor to use in a given application improves the cost, performance, and simplicity of your machine-design process. There are many different electrical motor types, all with their good and bad sides. Motion control is the art and science of precisely controlling the position, velocity, and torque of a mechanical drive. Motion-control systems comprise a numerical controller that performs path generation, such as a DSP; an amplifier; and a motor. Positioning-control systems most often employ step motors, dc-brush motors, and brushless-dc (permanent magnet) motors.

The control made on the electric motor is a position control, a speed control, and generally a torque control. The position might not be controlled when the main control is speed, the torque control is substituted by the current control. On the one side, the research of sensorless conversion was done actively in order to the reliable improvement and economical line and that conversions is demanded in the control where high accuracy is required. The speed sensorless vector control is the good example. There are not a lot of controls with a torque sensor because it achieved the torque control with $\pm 3\%$--$\pm 5\%$ error controlling the current, and when thrust control of $\leq 1\%$ or less error is required for the accuracy like the injection molding machine, the thrust torque is detected using the load cell. The thrust control system is set up in the speed control system and it is composed with the current control system inside. In this study, the torque control system must be precisely controlled thus the thrust conversion have adapted estimation of magnet flux which has temperature dependency that contributes to disturb the relationship between
current and torque which are directly related, therefore, for high accuracy of torque control, the torque is controlled without torque sensor which has some inherent problems with the parameters variations due to the environment effect.

1.2 STATUS OF SPMSM IN POWER ELECTRONIC TECHNOLOGY

Due to the popularity of the general purpose of surface permanent magnet synchronous motor, now the theoretic research on SPMSM is gradually going into depth. Moreover, the researchers have constant interests in improving the performance-cost ratio of the system using SPMSM. Considering the importance and complexity of power electronic technology, it is usually classified for study. For example, in Ref.[1], there is very importance interest for the pressure control which is scarcely used on machine tools. Nothing is as precious as the characteristic of this pressure control for injection molding machine, because it is directly traced to the quality of the molded components. For this reason, the development of the pressure control by a servomotor is indispensable for the achievement of the electrifier of the hydraulic injection molding machine to the injection molding machine with servomotors.

1.3 ADVANTAGE OF SENSORLESS CONTROL

For starting up the motor, three hall sensors can be distinguishing six commutation areas Ref.[2], [3]. Many motion control applications require the use of a position transducer for feedback, such as an encoder or a revolver. And for pressure control of SPMSM applications require the use of a magnet flux transducer for feedback to perform commutation. Some systems utilize velocity transducer as well. These sensors increase cost and weight but reduce the reliability of the system. Research in the area of sensorless control of the PMSM is beneficial because of the elimination of the feedback wiring, reduced cost, and improved reliability.

1.4 OVERVIEW OF TORQUE SENSORLESS CONTROL OF SPMSM

Recently, many plastic products have been produced and used. These plastic products are mostly manufactured using injection molding machines. The quality of plastic products depends on the injection force. Therefore, it is important to develop a fine force-control system. Generally, in force-control systems, the force information from the environment is detected by a force sensor. However, control systems using force sensors present problems related to signal noise, sensor cost,
narrow bandwidth, and other factors.
The quality of plastic products depends on the injection force. For that reason, it is important to develop not only a high-performance position control system but also a fine force-control system.

There has been a great deal of research in the area of torque sensorless control; we list them as follows:

A. Model-based torque control system using a torque observer Ref.[4]: This torque observer estimates the motor torque using motor current information and motor position.

B. The disturbance observer Ref.[5]: This compensates the disturbance torque for the motor. The disturbance observer estimates the disturbance torque without a torque sensor.

C. A robust tracking servo system for the optical disk recording system Ref.[6]: This system realizes a robust servo control using a force-sensorless method.

D. The sensorless force-control method using the reaction torque observer Ref.[7-9]: The reaction torque observer is based on the disturbance observer and friction model. This sensorless force-control method uses only the motor current information and motor position information. In other words, this torque estimation algorithm requires no additional sensor. A sensorless force control for an injection molding machine without any additional sensor has not been achieved yet. The reaction torque observer is applied to the injection molding machine using a ball screw. The motion control system using the ball screw often has a resonant frequency Ref.[10]–[12]. This torsional vibration affects the performance of reaction torque estimation.

In this thesis, the magnet flux estimation of vector control used adaptive identification. The observer response can be improved carefully because $d, q$ axis voltages are applied to both mathematical model and real machine SPMSM.

1.5 TYPES OF INJECTION MOLDING MACHINES

Machines are classified primarily by the type of driving systems they use: hydraulic, Mechanical, electric, or hybrid. Hydraulic presses have historically been the only option available to molders until Nissei Plastic Industrial Co., LTD introduced the first all-electric injection molding machine in 1983. The electric press, also known as Electric Machine Technology (EMT), reduces operation costs by cutting energy consumption and also addresses some of the environmental concerns surrounding the hydraulic press. Electric presses have been shown to be quieter, faster, and have a higher accuracy; however the machines are more expensive. Mechanical type
machines use the toggle system for building up tonnage on the clamp side of the machine. Tonnage is required on all machines so that the clamp side of the machine does not open (tool half mounted on the platen) due to the injection pressure. If the tool half opens up it will create flash in the plastic product. Reliability of mechanical type of machines is more as tonnage built during each cycle is the same as compared to hydraulic machines. Hybrid injection molding machines claim to take advantage of the best features of both hydraulic and electric systems, but in actuality use almost the same amount of electricity to operate as a standard hydraulic. Hydraulic machines, although not nearly as precise, are the predominant type in most of the world, with the exception of Japan. Nevertheless, all of our study focused on electric presses of injection molding machine that use surface permanent magnet synchronous motor.

Fig. 1-1. Topography of injection molding mechanism

Fig. 1-2. Schematic diagram of the injection molding machine using force sensor
1.6 ADVANCED CONTROL STRATEGY \(^{[13-15]}\)

The control system can improve characteristic of the pressure control for an injection molding machine. Some advanced control strategies are implemented since DSP can timely figure out the output of PWM according to the feedback value. Therefore, the detected current and voltage are fed to the input of DSP and then calculates the voltage order. 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. To achieve the fast dynamical response and high steady-state wave precision of output voltage of inverter, the advanced control strategies, such as voltage sensor, space vector control, current detector (Hall Current transformer), current sensor, are applied by the researchers in accordance with the special conditions and requirements. The response speed of the inverter is required to much faster than that of the motion system of motor in consideration of the obvious delay of digital control during fast switching.

The other subjects of the technique to be deeply investigated include: how to establish the mathematical model of the control objects and how to calculate the lumped equivalent parameters under working conditions.

1.7 COMPUTER SIMULATION

Computer simulation can promote the development process of SPMSM to a large extent. As a research method based on the vector control of SPMSM using adaptive identification method, the estimated magnet flux and armature winding resistance simulation is to build the component/system mathematical model and real SPMSM and to numerically analyze it on computer under the specific boundary conditions and time-history conditions and to finally visualize the results. The magnet flux and armature winding resistance simulation is verified how the current observer through the mathematical gains conduct the magnet flux and armature winding resistance estimation resulting from the vector control using adaptive control. Therefore, the large signal analysis is available for the switching the magnet flux and armature winding resistance estimation can be determined. In addition, as the general-purpose development tools of estimation of magnet flux and armature winding resistance in the technical and engineering fields, the commercial software of Matlab/simulink bring the convenience for the power electronic component and system simulations.

The other subjects of the technique to be deeply investigated include: how to
establish the mathematical model of the control objects and how to adjust the observer gains under working conditions.

1.8 NECESSITY

In general, ac drives are high order nonlinear systems. To apply linear system theory to controller design, the governing equations must be linearized. The output variable is measured when precision is required. Otherwise, the output is estimated through observers, when precision is not of a major concern. As the drives are dual input (torque, flux) and multistate variable (currents, voltages, speed, position, etc.) systems, the generally utilize state feed back. Some of the state variables are measured, and some are estimated from the measured ones, knowing the motor parameters and load. However, the motor parameters depend on temperature (resistance) and saturation (inductance). The load on the motor may also change. All these variations affect the closed-loop response in fixed structure fixed gain input-output (or state feedback) controllers. Increasing the gain of the controller results in robustness in response (less sensitivity to parameter and load variations), but disturbs the behavior in the region of reference torque, speed, or position. Three main methods have been proposed to circumvent this difficulty: (a) self-tuning controllers Ref.[16]; (b) reference model adaptive controllers Ref.[17]; (c) variable structure controllers Ref.[18].

When the number of variables is small, as in an induction machine, the variables are estimated through an observer. A self-tuning vector controller is thus obtained. For example, the value of the rotor resistance of an induction motor is estimated and updated in the vector controller, rather than changing the gains of the position and speed controllers. This approach yields robustness to motor parameter detuning and thus solves part of the problem. However, the method requires a considerable amount of computing time unless simplified configurations in the observer are used. The second part of problem is robustness to inertia and load disturbances. Model reference adaptive controllers (MRAC) have been proposed for this purpose Ref.[19]. In essence, an MRAC uses a reference model of the plant (converter-motor-load) dynamics with nominal parameters $P_m(s)$. The corresponding output is compared with the actual output. The out error thus obtained is fed forward in the control through a low-pass filter and $P_m^{-1}(s)$, where $P_m(s)$ represents the actual dynamics of the system.
1.9 PURPOSE AND SCOPE OF THE STUDY

The aim of this study is to remove the force sensor in injection molding machine and assure the constant torque (constant pressure) during the operation of the machine through the magnet flux estimation of vector control of surface permanent magnet synchronous motor using adaptive identification to overcome the temperature issue which affect the armature winding resistance and then the magnet flux Ref.[20] and robustness to inertia and load disturbances Ref.[21] Within this context, in order to realize the effectiveness of the proposed method, the estimators were designing through from linearization error of state equation in \((d,q)\) coordinates and the mathematical model of SPMSM.

In this regard, experimental studies would be carried out by using the reference model of plant (converter-motor-load) in order to evaluate the effectiveness of magnet flux and armature winding resistance estimation for torque (pressure) control during the injection molding machine operation. The results of multiple experimental Runs were presented along the literature of thesis based on magnet flux estimation at different torque and speed. All the experimental shows in the next chapter demonstrated the effectiveness of magnet flux estimation of SPMSM.

We give a brief overview of this thesis:

Chapter1, introduction, introduces the permanent magnet synchronous motor control system.

Chapter 2, research on torque control method of vector control of SPMSM.

Chapter 3, express the permanent magnet flux estimation method of vector control SPMSM using adaptive identification.

At first, it describes in detail with the experimental result the influences of parameters errors; the magnet flux and armature winding resistance are estimated through the currents errors using the both mathematical model and real machine equations.

Chapter4, Robustness. This chapter describe that the magnet flux can be estimated without the influences of armature winding resistance.

Chapter5, the general conclusion
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CHAPTER 2

TORQUE CONTROL METHOD OF VECTOR CONTROL OF SPMSM

2.1 INTRODUCTION

Surface magnet type permanent magnet synchronous electric motor (SPMSM: Surface Permanent Magnet Synchronous Motor) is widely used in industry through its advantages: to be small, high-power density, high efficiency, and easy to control. Among the electrical system of machines, the SPMSM are frequent in industry and this thesis study the control system of them. There are various uses of the motor in industry and the better economics of them is the SPMSM. In industry, injection molding machine uses the SPMSM drive. This research is thought to overtake the power sensor less conversion in order to further the economical efficiency because the piezoelectric device used in the power sensor which has dependency to the temperature and amends its performance. For sensor less, it is possible to know this power namely the torque through the magnet flux of SPMSM and the compensation of the power sensor and temperature related to that become unnecessary. However, it is well known that the SPMSM motors’ constant-power speed range during flux-weakening operation has been limited due to relatively large air-gap.

In the normal range of operating temperature, as the temperature increases, the residual flux density and intrinsic coercivity of the magnet will decrease Ref. [1],[2]. This is a reversible process that when the temperature decreases and the flux density and coercivity will return to its original value Ref.[3]. The variation in residual flux density of the magnet, along with the variation in armature resistance of the motor with temperature, influences the torque capability and power efficiency of the PM motor. When the operating temperature of the magnet increases above a critical temperature, it will result in irreversible demagnetization of the magnet. Once this happens, the flux density will not go back to its original value as the temperature decreases. The critical temperature at which point the irreversible demagnetization occurs is a function of the magnet and the load operating curve of
the magnet. A high operation temperature of an armature winding current impulse may cause magnet flux ripple or demagnetization, which has direct impact on machine performance Ref.[4], [5]. Regarding to demagnetization, much research has been undertaken Ref.[6-12]. This study expresses also, the torque, speed controller, current controller from the base to the software servo.

2.2 CIRCUIT AND MOTION EQUATION OF MOTOR\textsuperscript{[13]}

When the motor is controlled, the circuit equation is a base for the grasp of the control characteristic and deriving the control scheme. Here, is described the circuit equation of the induction motor where the permanent magnet synchronous motor used as brushless DC motor and the inverter is driven. Moreover, the motion equation is described.

2.2.1 Structure

A permanent magnet synchronous motor is a revolving-field type where the field magnet rotates to assume the brushless with a synchronous motor which uses a permanent magnet for the field magnet. Figure 2.1 shows the structure of cylindrical shape.

![Fig. 2-1 Structure of surface permanent magnet synchronous motor](image)

2.2.2 Representation of three-phase AC circuit equation

Figure 2-2 is the equivalent circuit of 3 phase cylindrical permanent magnet synchronous motor. The circuit equation of the relation of voltage, current and impedance from the equivalent circuit becomes.
\[
\begin{bmatrix}
  v_{ua} \\
  v_{vb} \\
  v_{wc}
\end{bmatrix} =
\begin{bmatrix}
  R_a + PL_a & -\frac{1}{2} PM_a' & -\frac{1}{2} PM_a' \\
  -\frac{1}{2} PM_a' & R_a + PL_a & -\frac{1}{2} PM_a' \\
  -\frac{1}{2} PM_a' & -\frac{1}{2} PM_a' & R_a + PL_a
\end{bmatrix}
\begin{bmatrix}
  i_{ua} \\
  i_{vb} \\
  i_{wc}
\end{bmatrix} +
\begin{bmatrix}
  e_{ua} \\
  e_{vb} \\
  e_{wc}
\end{bmatrix}
\]

(2.1)

Here, \( v_{ua}, v_{vb}, v_{wc} \) are the armature voltage of \( u, v, w \) phase. \( i_{ua}, i_{vb}, i_{wc} \) are the armature current of \( u, v, w \) phase. \( e_{ua}, e_{vb}, e_{wc} \) are a speed electromotive force which is induced in the \( u, v, w \) phase armature winding by permanent magnet magnetic field. \( R_a \) is the armature winding resistance. \( L'_a \) is self-inductance of the armature winding. \( M'_a \) is a mutual inductance between armature winding. \( P(= d/dt) \) is a differential operator.

When the maximum value is assumed to be \( \Phi_{fa}, \Phi_{fb}, \Phi_{fc}, \Phi_{fw} \) field magnets of \( u, v, w \) phase armature winding interlinkage fluxes to generate \( e_{ua}, e_{vb}, e_{wc} \) becomes.

**Fig. 2-2** The equivalent circuit of three-phase cylindrical permanent magnet synchronous motor
Here, $\theta_{re}$ is an angle of the field magnet taken clockwise based on $u$ phase of armature winding (electrical angle), and $\omega_{re}$ the angular velocity of the magnetic field (electrical angle) are express as follow:

$$\theta_{re} = \int \omega_{re} dt$$  \hspace{1cm} (2.3)

In this case, $e_{ua}, e_{va}, e_{wa}$ becomes:

$$e_{ua} = P\phi_{fua} = -\omega_{re} \Phi'_{re} \sin \theta_{re}$$
$$e_{va} = P\phi_{fva} = -\omega_{re} \Phi'_{re} \sin \left( \theta_{re} - \frac{2\pi}{3} \right)$$  \hspace{1cm} (2.4)
$$e_{wa} = P\phi_{fwu} = -\omega_{re} \Phi'_{re} \sin \left( \theta_{re} + \frac{2\pi}{3} \right)$$

In the armature winding, there is also a leakage inductance $l_a$ and its relation with the self-inductance of the armature winding is express in the next equation.

$$L'_a = l_a + M'_a$$  \hspace{1cm} (2.5)

Furthermore, the number of pole pairs is assumed to be $p$, the rotation speed $\omega_{rm}$ of the output shaft of a synchronous motor (mechanical angle) is $\omega_{re} / p$.

### 2.2.3 Coordinate transformation

As for the grasp of the control system characteristic deriving the control method, it is easier when it is represented by 2 phase than to be represented by 3 phase alternative current and voltage. Moreover, it is simple to represent 2 axis direct current than 2 phase alternative current. To change the view of the motor this way, it is necessary to change the coordinate view, this is called coordinate transformation.

### 2.2.4 Two phase circuit equation

#### 2.2.4.1 ($\alpha - \beta$) circuit equation coordinate system

The circuit equation of ($\alpha - \beta$) from 3 phase alternative current through the
coordinate transformation is shown in equation (2.6).

\[
\begin{bmatrix}
    v_{\alpha\alpha} \\
v_{\beta\alpha}
\end{bmatrix} =
\begin{bmatrix}
    R_a + PL_a & 0 \\
0 & R_a + PL_a
\end{bmatrix}
\begin{bmatrix}
i_{\alpha\alpha} \\
i_{\beta\alpha}
\end{bmatrix} +
\begin{bmatrix}
e_{\alpha\alpha} \\
e_{\beta\alpha}
\end{bmatrix}
\]  

(2.6)

Figure 2.4 shows the equivalent circuit.

Fig. 2-3 The equivalent circuit 2 phase alternative current

Here, \( v_{\alpha\alpha}, v_{\beta\alpha} \) are \( \alpha, \beta \) axis armature voltage, \( i_{\alpha\alpha}, i_{\beta\alpha} \) are \( \alpha, \beta \) phase armature current, \( e_{\alpha\alpha}, e_{\beta\alpha} \) are the speed electromotive force induced by \( \alpha, \beta \) phase armature windings of the field permanent magnet. \( R_a \) is armature winding resistance, \( L_a \) is self-inductance of the armature winding, \( R_a \) is the same as equation (2.1). \( L_a \) is represented in the next equation using \( I_a, M_a' \) of equation (2.1).

\[
L_a = I_a + \frac{3}{2} M_a'
\]  

(2.7)

2.2.4.2 \( d-q \) circuit equation coordinate system

The motor has the fixed and rotating parts. Converting them into the orthogonal coordinate system where orthogonal coordinate system \( d-q \) transformation rotates them both fixed, that system of coordinates is \( d-q \) coordinate system. \( q \)-axis has \( \pi/2 \) phase advanced compared to \( d \)-axis.

The circuit equation of \( d-q \) from \( (\alpha - \beta) \) circuit equation coordinate system is:

\[
\begin{bmatrix}
    v_{d\alpha} \\
v_{q\alpha}
\end{bmatrix} =
\begin{bmatrix}
    R_a + PL_a & -\omega R_a L_a \\
\omega R_a L_a & R_a + PL_a
\end{bmatrix}
\begin{bmatrix}
i_{d\alpha} \\
i_{q\alpha}
\end{bmatrix} +
\begin{bmatrix}
e_{d\alpha} \\
e_{q\alpha}
\end{bmatrix}
\]  

(2.8)

The second term of right side of this equation generated speed electromotive force \( e_{d\alpha}, e_{q\alpha} \) in \( d-q \) axis of armature winding by the permanent magnet magnetic field.
are \( e_{d0} = 0, e_{q0} = \omega_e \phi_{f0} \).

If \( v_{d0}, v_{q0} \) are assumed to be a direct voltage, \( i_{d0}, i_{q0} \) becomes direct current too and it is possible to treat in two axis direct current. Furthermore, because the field magnet is in \( d \)-axis, it is generated only in an advanced on \( q \)-axis of \( \pi/2 \), and the right term of equation (2.8) is a direct voltage in \( d \)-\( q \) axis armature winding according to the field magnetic of permanent magnet as described earlier and generates speed electromotive force.

### 2.2.5 Torque

\( T_e \) is the motor torque generated by the Fleming’s left hand law. However, the revolving-field type motor torque is the torque applied to the field (Fleming’s left hand law in the positive direction of torque has been applied to the armature windings). It is represented by the sum of product of orthogonal armature current and armature winding flux. Torque equation is:

\[
T_e = p \phi_{fa} \left\{ -i_{u0} \sin \theta_{re} - i_{u0} \sin \left( \theta_{re} - \frac{2\pi}{3} \right) - i_{u0} \sin \left( \theta_{re} + \frac{2\pi}{3} \right) \right\} \\
= p \phi_{fa} \left\{ -i_{u0} \sin \theta_{re} + i_{fa} \cos \theta_{re} \right\} \\
= p \phi_{fa} i_{qu} \quad (2.9)
\]

### 2.3 Decoupling control law

A permanent magnet synchronous motor has the speed electromotive force that
interferes between d-q axes mutually. The direct regulation can not be done though they influence $i_{da}, i_{qa}$. Then, the speed electromotive force is requested and it is thought the control that denies it. That is a decoupling control, and $v_{da}, v_{qa}$ are controlled with:

$$
\begin{align*}
    v_{da} &= v_{da}' - \omega_r L_a i_{qa} \\
    v_{qa} &= v_{qa}' + e_{qa} + \omega_r L_a i_{da} \\
    &= v_{qa}' + \omega_r (\phi_{fa} + L_a i_{da})
\end{align*}
$$

(2.10)

$\omega_r, i_{da}, i_{qa}$ can be detected though neither speed electromotive force $\omega_r L_a i_{qa}$ nor $\omega_r (\phi_{fa} + L_a i_{da})$ can directly control. Because $L_a, \phi_{fa}$ are constant that it can be measured beforehand, it requests and operates in the control circuit. $i_{da}, i_{qa}$ are obtained from $i_{wa}, i_{wa}$ by a coordinate transformation. Consequently, the circuit becomes:

$$
\begin{bmatrix}
    v_{da}' \\
    v_{qa}'
\end{bmatrix}
= \begin{bmatrix}
    R_a + PL_a & 0 \\
    0 & R_a + PL_a
\end{bmatrix}
\begin{bmatrix}
    i_{da} \\
    i_{qa}
\end{bmatrix}
$$

(2.11)

It is understood that, knowing $v_{da}', v_{qa}'$, $i_{da}, i_{qa}$ can be controlled simply as expressed by state equation.

$$
P \begin{bmatrix}
    i_{da} \\
    i_{qa}
\end{bmatrix}
= \begin{bmatrix}
    R_a & 0 \\
    0 & R_a
\end{bmatrix}
\begin{bmatrix}
    i_{da} \\
    i_{qa}
\end{bmatrix}
+ \begin{bmatrix}
    1 & 0 \\
    0 & 1
\end{bmatrix}
\begin{bmatrix}
    v_{da}' \\
    v_{qa}'
\end{bmatrix}
$$

(2.12)

The equation (2.11) shows the armature winding impedance applied by $v_{da}', v_{qa}'$ d-q axis voltage, and it becomes an input variable able to control the decoupling state.

### 2.4 Entire configuration of vector control system

SPMSM is the controlled system, voltage type PWM inverter, including the encoder for the pole position detection etc..., shows the entire composition of vector control in the figure 2-5.
Fig. 2-5 Entire compositions of d-q coordinate of brushless DC motor control

The figure 2-5 shows the entire composition of d-q coordinate of brushless DC motor control. It is observed that, the output of speed controller is the torque reference current $i_{qa}^*$. 

2.5 DEMAGNETIZATION CHARACTERISTICS

Fig. 2-6 Typical demagnetization characteristics of common magnets

Rare earth permanent magnets have emerged as the key components of the high-tech electro-mechanical devices. It is simply due to the fact the utilization of rare earth magnets can make the devices much more energy efficient, light weight, and compact. High performance rare earth magnets use rare earths as their main
constituents. There are two major series of rare earth magnets commercially available which are SmCo and NdFeB. The common features of these magnetic materials are their high remanence $B_r$ (remanent flux density) and high coercivity $H_c$ (coercive force) compared with those of conventional magnets (AlNiCo, ferrite). In addition, the demagnetization curve is a straight line in the second quadrant of the B–H curve. Therefore, this type of magnet is very resistive to demagnetization fields. In Figure 2-6, shown are the typical demagnetization characteristics of the rare earth magnets along with those of conventional ones.

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 has the dependency in 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%/°C. When the temperature of the motor changes between ±75° C, the variation of the permanent magnet flux is within ±9% and the armature winding resistance changes is 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.

2.6 CONCLUSION

It observed in this chapter that, the torque can not be constant during the operation after some time because of the temperature which increase in the motor (SPMSM) causing the increase the armature winding resistance which cause the decrease of magnet flux. And, since the magnet flux is related to the torque then the torque will decrease as well. The next chapter expresses the estimation of the magnet flux and the armature winding resistance using adaptive identification in order to improve the accuracy of control of Surface permanent magnet synchronous motor without torque sensor.
REFERENCES


CHAPTER 3

PERMANENT MAGNET FLUX ESTIMATION METHOD OF VECTOR CONTROL SPMSM USING ADAPTIVE IDENTIFICATION

3.1 INTRODUCTION

Permanent magnet synchronous motor (PMSM) has been receiving much attention because of the inherent advantages of high-power density and high efficiency. As one of the rare-earth family of permanent magnets, the neodymium-iron-boron (Nd-Fe-B) material is often referred as one of the most advanced permanent magnet material available today. Because of its high magnet flux density and lower cost than the other rare-earth materials, it allows a small size with high magnetic fields application. Due to its high efficiency, high torque density, and lower cogging torque features Ref.[1], [2], PMSM with Nd-Fe-B material has attracted increasing interests among researchers and designers in many high performance applications such as industrial servo systems. However, the applications and operational environments impacts to the permanent magnet (PM) materials have seldom been evaluated Ref.[3]

Several methods have been proposed to improve the performance of a PMSM drive by estimating the electrical parameters Ref.[4-15].

Recently, plastic has become the most widely used raw material in various fields. Plastic products are mainly manufactured using injection molding machines. Many studies of electric injection molding machines have been carried out Ref.[16–20]. High-performance position control enables large-scale production of plastic products. However, the quality of plastic products depends on the injection force. For that reason, it is important to develop not only a high-performance position control system but also a fine force-control system. Regarding conventional force control, much research has been undertaken to develop force sensors to detect external force Ref.[21–23]. A typical injection molding machine senses the force information using a force sensor.
However, highly sensitive force sensors are not economical. A typical force sensor has both initial and running costs. Moreover, force sensors confront problems such as noise and frequency bands. In an ideal force-control system, force sensors should be attached to the same location as the actuator to realize an instantaneous force sensing process. However, in a conventional actual servo system, force sensors are mounted on different positions than the actuator. Consequently, it is difficult for force sensing systems to obtain force data accurately and instantaneously. To overcome these problems, many force-sensorless control methods have been proposed. Sun and Mills described a model-based torque control system using a torque observer Ref.[24]. This torque observer estimates the motor torque using motor current information and motor position. Ohnishi et al, proposed the disturbance observer, which compensates the disturbance torque for the motor Ref.[25]. The disturbance observer estimates well the disturbance torque without a torque sensor. Ohishi et al, proposed a robust tracking servo system for the optical disk recording system Ref.[26]. This system realizes a robust servo control using a force-sensorless method. Furthermore, the sensorless force-control method using the reaction torque observer has been applied Ref.[27–29]. The reaction torque observer is based on the disturbance observer and friction model. This sensorless force-control method uses only the motor current information and motor position information. In other words, this torque estimation algorithm requires no additional sensor. A sensorless force control for an injection molding machine without any additional sensor has not been achieved yet. The reaction torque observer is applied to the injection molding machine using a ball screw. The motion control system using the ball screw often has a resonant frequency Ref.[30–32]. This torsional vibration affects the performance of reaction torque estimation. It has proposed also reaction torque observer based on a two-inertia plant model considering the torsion phenomenon Ref.[33],[34]. To overcome this problem, this study presents a magnet flux estimation system in order to control torque without torque sensor (Force sensor or load cell) controlling armature current with error less than 1% by using adaptive identification. The salient features of the propose methods are showed 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, deriving the linearization error state equation of a real machine and using the 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 research, the estimator is composed by linearization error margin of state equation from state equation of a real machine and the mathematical model. The utility is confirmed by both of the simulations and the experiments.

### 3.2 COMPOSITION OF TORQUE REFERENCE

The Figure 3-8 shows the composition of the proposed system. The instruction value is shown with * in this research. The voltages \( v_{da}, v_{qa} \) are given to a real SPMSM and mathematical model as an input. From which the mathematical model estimates \( \hat{\phi}_{fa}, \hat{R}_a \) the quantities \( \Delta \hat{\phi}_{fa}, \Delta \hat{R}_a \) are causes of the current estimation error and these may become 0 by using the estimation of the output current.

Usually, the output of the speed controller is the torque reference. However, the particularity of this method is that the torque reference current \( i_{qa}^* \) is obtained by computation that used the output of speed controller and the estimated magnet flux. Therefore, the estimated parameter is used in controller. The torque expression is showed in equation (3.1) and reference current construction is showed in the Figure 3-1.

\[
\hat{T}_e = p \hat{\phi}_{fa} i_{qa} \tag{3.1}
\]

\( \hat{T}_e \) : estimated torque

\( p \) : number of pole pairs

\( \hat{\phi}_{fa} \) : estimated magnet flux

\( i_{qa} \) : q axis current
3.3 MATHEMATICAL MODEL AND LINEARIZATION ERROR EQUATION OF STATE

The adaptive identification is to construct the observer based on the reference model of the controlled system of plant (converter-motor-load) and make for an unknown plant parameter. The controlled system is SPMSM, and in the proposed vector control of the magnet flux estimation method, the unknown parameters are magnet flux and armature winding resistance. The estimation function of the state variable and the identification function of the unknown parameters are both in the observer. It is described in this chapter and next chapter the estimator of the mathematical model. The proposed method described the estimator d,q axis armature current which is the state variable according to the mathematical model, and identifies the magnet flux and the armature winding resistance that are the unknown parameters in the estimator.

In this work, the magnet flux is estimated by using the mathematical model and also the estimation error of state equation is obtained through the state equation of a real and the mathematical model of SPMSM.

3.3.1 Mathematical model

The state equation of SPMSM in (d, q) coordinates is shown in equation (3.2). Moreover, the equation (3.3) shows the mathematical model in Figure 3-10 by which the estimated current is described, $\hat{i}_{da}$, $\hat{i}_{qa}$ are estimated by using

$i_{da}, i_{qa}, v_{da}, v_{qa}, \phi_{fa}, R_a$
\[
\begin{bmatrix}
L_a & 0 \\
0 & L_a
\end{bmatrix}
\frac{d}{dt}
\begin{bmatrix}
i_{da} \\
i_{qa}
\end{bmatrix}
= \begin{bmatrix}
-R_a & \omega L_a \\
-\omega L_a & -R_a
\end{bmatrix}
\begin{bmatrix}
i_{da} \\
i_{qa}
\end{bmatrix}
+ \begin{bmatrix}
v_{da} \\
v_{qa}
\end{bmatrix}
- \omega \phi_{fa}
\begin{bmatrix}
0 \\
1
\end{bmatrix}
\]
(3.2)

\[
\begin{bmatrix}
L_a & 0 \\
0 & L_a
\end{bmatrix}
\frac{d}{dt}
\begin{bmatrix}
i_{da}^\wedge \\
i_{qa}^\wedge
\end{bmatrix}
= \begin{bmatrix}
-R_a & \alpha L_a \\
-\alpha L_a & -R_a
\end{bmatrix}
\begin{bmatrix}
i_{da}^\wedge \\
i_{qa}^\wedge
\end{bmatrix}
+ \begin{bmatrix}
v_{da}^\wedge \\
v_{qa}^\wedge
\end{bmatrix}
- \omega \phi_{fa}^\wedge
\begin{bmatrix}
0 \\
1
\end{bmatrix}
+ \begin{bmatrix}
g_{11} & g_{12} \\
g_{21} & g_{22}
\end{bmatrix}
\begin{bmatrix}
i_{da} - i_{da}^\wedge \\
i_{qa} - i_{qa}^\wedge
\end{bmatrix}
\]
(3.3)

\(R_a\): armature winding resistance, \(L_a\): inductance, \(\phi_{fa}\) permanent magnet flux,

\(v_{da}, v_{qa}\): d,q axes voltages, \(i_{da}, i_{qa}\): d,q axes currents, \(\omega_{re}\): synchronous angular frequency.

In this work, the current of the armature and the parameter with\(^\wedge\) shows “estimation” and then the estimated parameters are \(\phi_{fa}, R_a\) Moreover,

\(g_{11}, g_{12}, g_{21}, g_{22}\) are mathematical model gains.

### 3.3.1.1 The estimation error of state equation

The estimation error of state equation is requested from the difference of the state equation of a real and the mathematical model of SPMSM. To proceed, we consider the equations (3.2) and (3.3).

The equation (3.2) can be written as follows:

The first line is:

\[
\frac{d}{dt} i_{da} = \frac{1}{L_a} \left( -R_a i_{da} + \omega_{re} L_a i_{qa} + v_{da} \right)
\]
(3.4)

And the second line is:

\[
\frac{d}{dt} i_{qa} = \frac{1}{L_a} \left( -\omega_{re} L_a i_{da} - R_a i_{qa} + v_{qa} - \omega_{re} \phi_f \right)
\]
(3.5)

And the equation (3.3) is written as follows:

The first line is:
\[
\frac{d}{dt} i_{da}^\wedge = \frac{1}{L_a} \left( -R_a i_{da}^\wedge + \omega_{re} L_a i_{qa}^\wedge + v_{da} \right) + g_{11} i_{da} - i_{da}^\wedge) + g_{12} \left( i_{qa} - i_{qa}^\wedge \right)
\]

(3.6)

And the second line is:

\[
\frac{d}{dt} i_{qa}^\wedge = \frac{1}{L_a} \left( -\omega_{re} L_a i_{da} - R_a i_{qa} + \omega_{re} \phi_f \right) - g_{21} \left( i_{da} - i_{da}^\wedge) + g_{22} \left( i_{qa} - i_{qa}^\wedge \right)
\]

(3.7)

The difference between the equations (3.2) and (3.3) is:

\[
\frac{d}{dt} \left[ i_{da} - i_{da}^\wedge \right] = \frac{1}{L_a L_a} \left[ L_a \begin{array}{c} L_a \end{array} \begin{array}{c} 0 \end{array} \begin{array}{c} L_a \end{array} \begin{array}{c} \omega_{re} L_a i_{qa} + \phi_r \end{array} \end{array} \left[ -R_a i_{da} - \omega_{re} L_a i_{qa} + \phi_r \right] \right] - \omega_{re} \phi_r \left[ \begin{array}{c} 0 \\ 1 \end{array} \right] = \omega_{re} \phi_r \left[ \begin{array}{c} 0 \\ 1 \end{array} \right]
\]

26
\[
\begin{align*}
&= \frac{1}{L_a L_a} \begin{bmatrix} L_a & 0 \\ 0 & L_a \end{bmatrix} \begin{bmatrix} -i_{da} \omega \omega \\ -i_{qa} \end{bmatrix} \begin{bmatrix} \hat{R}_a \omega \\ -\hat{R}_a \omega \end{bmatrix} + \begin{bmatrix} \omega_r L_a - g_{12} \\ -\omega_r L_a + g_{21} \end{bmatrix} \begin{bmatrix} i_{da} - i_{da} \omega \\ i_{qa} - i_{qa} \end{bmatrix} - \omega_r \phi_{\mathbf{f}_a} - \phi_{\mathbf{f}_a} \begin{bmatrix} 0 \\ 1 \end{bmatrix} \\
&= \frac{1}{L_a L_a} \begin{bmatrix} L_a & 0 \\ 0 & L_a \end{bmatrix} \begin{bmatrix} -i_{da} \Delta R_a \\ -\omega_r L_a + g_{21} \\ -\omega_r L_a + g_{21} \end{bmatrix} \begin{bmatrix} i_{da} - i_{da} \omega \\ i_{qa} - i_{qa} \end{bmatrix} + \begin{bmatrix} \omega_r \Delta \phi_{\mathbf{f}_a} \\ -\omega_r \Delta \phi_{\mathbf{f}_a} \end{bmatrix}
\end{align*}
\]

Therefore, the estimation error of state equation is:

\[
\begin{align*}
\mathbf{s} = \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} -\hat{R}_a - g_{11} \omega _{\mathbf{r}} L_a - g_{12} \\ -\omega _{\mathbf{r}} L_a + g_{21} \hat{R}_a - g_{22} \\ -\omega _{\mathbf{r}} L_a + g_{21} \hat{R}_a - g_{22} \end{bmatrix} & \begin{bmatrix} i_{da} - \hat{i}_{da} \omega \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} 0 \\ -\omega _{\mathbf{r}} \Delta \phi_{\mathbf{f}_a} \end{bmatrix} \begin{bmatrix} \Delta \phi_{\mathbf{f}_a} \\ \Delta \phi_{\mathbf{f}_a} \end{bmatrix} 
\end{align*}
\]

(3.8)

### 3.3.1.2 Diagonal of axis

For diagonal, the non interference of axis is necessary and to do it, \( v_{da}, v_{qa} \) are settled:

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix} i_{da} \\ i_{qa} \end{bmatrix} &= \begin{bmatrix} \frac{-R_a}{L_a} & \omega \\ -\omega & \frac{-R_a}{L_a} \end{bmatrix} \begin{bmatrix} i_{da} \\ i_{qa} \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} v_{da} \\ v_{qa} \end{bmatrix} - \omega \phi_{\mathbf{f}_a} \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \phi_{\mathbf{f}_a} \\ \phi_{\mathbf{f}_a} \end{bmatrix}
\end{align*}
\]

(3.9)

\[
\omega \times i_{qa} + \frac{1}{L_a} v_{da} = \frac{1}{L_a} \nu_{da}
\]

(3.10)

\[
\nu_{da} = v_{da} + \omega L_a i_{qa} \quad (v_{da} = v_{da} - \omega L_a i_{qa})
\]

(3.11)

\[
-\omega \times i_{da} + \frac{1}{L_a} v_{qa} - \omega \phi_{\mathbf{f}_a} = \frac{1}{L_a} \nu_{qa}
\]

(3.12)

\[
v_{qa} = v_{qa} - \omega L_a i_{da} - \omega \phi_{\mathbf{f}_a} \quad (v_{qa} = v_{qa} + \omega L_a i_{da} + \omega \phi_{\mathbf{f}_a})
\]

(3.13)
\[
\frac{d}{dt}\begin{bmatrix} i_{da} \\ i_{qa} \end{bmatrix} = \begin{bmatrix} -\frac{R_a}{L_a} & 0 \\ 0 & -\frac{R_a}{L_a} \end{bmatrix}\begin{bmatrix} i_{da} \\ i_{qa} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_a} & 0 \\ 0 & \frac{1}{L_a} \end{bmatrix}\begin{bmatrix} v'_{da} \\ v'_{qa} \end{bmatrix}
\]

(3.14)

\[\omega L_a i_qa\]

\[i_{da}^* \quad + \quad \text{----} \quad k_i \left( s + \frac{R_a}{L_s} \right) \quad v'_{da} \quad \text{----} \quad v_{da}\]

\[i_{da} \quad \text{----} \quad - \quad \text{----}\]

**Fig. 3-2  The structure of axis**

### 3.3.2 Linearization of estimation error of state equation

The equation (3.8) is non-linear and it is necessary to make it linear. The first approximates is divided into the equilibrium point and variation. The result of linearization of estimation error of state equation is derived as follows.

\[
S \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} -\hat{R}_a - g_{11} L_a & \omega_r L_a - g_{12} \\ -\omega_r L_a + g_{21} L_a & -\hat{R}_a - g_{22} L_a \end{bmatrix}\begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} 0 & \frac{i_{da}}{L_a} \\ \omega_r - \frac{i_{qa}}{L_a} \end{bmatrix}\begin{bmatrix} \Delta \phi_{fa} \\ \Delta R_a \end{bmatrix}
\]

(3.15)

Here, the mathematical model gains are \( g_{12} = \omega_r L_a \), \( g_{21} = -\omega_r L_a \)

\( g_{11} = g_{22} = g_a \) and equation (3.15) becomes:

\[
S \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} -\hat{R}_a + g_a L_a & 0 \\ 0 & -\hat{R}_a + g_a L_a \end{bmatrix}\begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} 0 & \frac{i_{da}}{L_a} \\ \omega_r - \frac{i_{qa}}{L_a} \end{bmatrix}\begin{bmatrix} \Delta \phi_{fa} \\ \Delta R_a \end{bmatrix}
\]

(3.16)

This equation can be rewritten as follows:
\[
\frac{d}{dt}\begin{bmatrix}
  i_{da} - \hat{i}_{da} \\
  i_{qa} - \hat{i}_{qa}
\end{bmatrix}
= \begin{bmatrix}
  A_{11} & A_{12} \\
  A_{21} & A_{22}
\end{bmatrix}
\begin{bmatrix}
  i_{da} - \hat{i}_{da} \\
  i_{qa} - \hat{i}_{qa}
\end{bmatrix}
+ \begin{bmatrix}
  B_{11} & B_{12} \\
  B_{21} & B_{22}
\end{bmatrix}
\begin{bmatrix}
  \Delta \phi_f \\
  \Delta R_a
\end{bmatrix}
\]

\[
= A \ e_{ia} + B \ u
\]  

(3.17)

Where:

\[
A_{11} = -\left( \hat{R}_a + g_{11} \right)/L_a, A_{12} = (\omega_r L_a - g_{12})/L_a, A_{21} = (-\omega_r L_a + g_{21})/L_a
\]

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

\[
e_{ia} = \begin{bmatrix}
  i_{da} - \hat{i}_{da} \\
  i_{qa} - \hat{i}_{qa}
\end{bmatrix}, u = \begin{bmatrix}
  \Delta \phi_f \\
  \Delta R_a
\end{bmatrix}
\]

The above linear equation is consisting of the armature current, the armature voltage, the angular velocity and the armature winding resistance. The linearization makes the first separate approximation of equilibrium point (hereafter, it is shown that subscript 0 with lower right is an equilibrium point) and changes mathematical model gain. Each equilibrium point of the equations (3.2) and (3.3) are assumed to take the same value.

\[
s\begin{bmatrix}
  i_{da} - \hat{i}_{da} \\
  i_{qa} - \hat{i}_{qa}
\end{bmatrix}
= \begin{bmatrix}
  A_{110} & A_{120} \\
  A_{210} & A_{220}
\end{bmatrix}
\begin{bmatrix}
  i_{da} - \hat{i}_{da} \\
  i_{qa} - \hat{i}_{qa}
\end{bmatrix}
+ \begin{bmatrix}
  B_{110} & B_{120} \\
  B_{210} & B_{220}
\end{bmatrix}
\begin{bmatrix}
  \Delta \phi_f \\
  \Delta R_a
\end{bmatrix}
\]

\[
= A_0 e_{ia} + B_0 u
\]  

(3.18)

**3.3.3 Linearization of estimation error transfer function matrix**

The linearization of estimation error of state equation is written like equation (3.18) and by deriving it, the linearization of estimation error transfer function is obtained.

The linearization of estimation error transfer function is the one to assume \( u \) to be an input and \( e_{ia} \) an output.
The following equations are transformed:
\[
\begin{align*}
    s \varepsilon_{\omega} &= A_0 \varepsilon_{\omega} + B_0 u \\
    (sI - A_0) \varepsilon_{\omega} &= B_0 u \\
    \therefore \quad \varepsilon_{\omega} &= (sI - A_0)^{-1} \cdot B_0 u 
\end{align*}
\]
(3.19)

Through the coefficient matrix \(u\) in the equation (3.19), the linearization of estimation error transfer function \(P_0(s)\) is computed as follows:

\[
P_0(s) = (sI - A_0)^{-1} \cdot B_0 = \left[ \begin{array}{c}
    s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    0 \\
    0 \\
    \end{array} \right]^{-1}
\left[ \begin{array}{c}
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]
\left[ \begin{array}{c}
    0 \\
    -i_{da} \\
    -i_{wa} \\
    \end{array} \right]
\]

\[
= \left[ \begin{array}{c}
    1 \\
    0 \\
    0 \\
    \end{array} \right]
\left[ \begin{array}{c}
    s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]^{-1}
\left[ \begin{array}{c}
    0 \\
    -i_{da} \\
    -i_{wa} \\
    \end{array} \right]
\]

\[
= \left[ \begin{array}{c}
    1 \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]
\left[ \begin{array}{c}
    s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]^{-1}
\left[ \begin{array}{c}
    0 \\
    -i_{da} \\
    -i_{wa} \\
    \end{array} \right]
\]

\[
= \left[ \begin{array}{c}
    1 \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]
\left[ \begin{array}{c}
    s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]^{-1}
\left[ \begin{array}{c}
    -i_{da} \\
    -i_{wa} \\
    \end{array} \right]
\]

\[
And,
\[
P_0(s) = \left[ \begin{array}{c}
    1 \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]
\left[ \begin{array}{c}
    s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    0 \\
    \frac{\hat{R}_{a0} + g_{a0}}{L_a} \\
    \end{array} \right]^{-1}
\left[ \begin{array}{c}
    -i_{da} \\
    -i_{wa} \\
    \end{array} \right]
\]

\[
(3.20)
\]

Next chapter shows the composition of the proposed system using the matrix.
transfer function matrix $P_0(s)$ of linearization of estimation error to compute the matrix inverse $P_0^{-1}(s)$ for the structure of the proposed system.

3.4. COMPOSITION OF THE PROPOSED SYSTEM

In this chapter, it is proposed the matrix inverse of the dynamics equation of system of the magnet flux derived in section 3.3 using the linearization of estimation error transfer function and the estimator. Moreover, the stability of the estimation system is examined.

3.4.1 Derivation of reverse-transfer function matrix

Matrix inverse $P_0^{-1}(s)$ of transfer function matrix $P_0(s)$ of linearization of estimation error that is showed in equation (3.21) is derived as follows:

$$\begin{align*}
P_0^{-1}(s) &= \begin{bmatrix}
0 & \frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left( -\frac{\omega_{c0}}{L_a} \right)
\frac{1}{s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}} \left( -\frac{i_{da0}}{L_a} \right)
\end{bmatrix}^{-1} \\
&= \begin{bmatrix}
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\end{bmatrix}
\begin{bmatrix}
\frac{L_a i_{qa0}}{\omega_{c0} i_{da0}}
\frac{L_a}{i_{da0}}
0
\end{bmatrix}
\begin{bmatrix}
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\end{bmatrix}
\begin{bmatrix}
\frac{L_a}{\omega_{c0}}
0
\end{bmatrix}
\begin{bmatrix}
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)
\end{bmatrix}
\begin{bmatrix}
\frac{L_a}{\omega_{c0}}
0
\end{bmatrix}
(3.21)
\end{align*}$$

When integration is made to act on $P_0^{-1}(s)$, each element is shown as

$$(P_0^{-1})_{ij} \text{ : It means i line j row element of } P_0^{-1}(s).$$
\[
\frac{k}{s} (p^{-1})_{11} = \left(1 + \frac{g_{11}}{s}\right) \frac{k L_d i_m}{\omega i_{dm}}
\]

\[
g_{11} = \frac{R_a + g_{11}}{L_d}
\]

(3.22)

\[
\frac{k}{s} (p^{-1})_{22} = \left(1 + \frac{g_{22}}{s}\right) \left(1 - \frac{k L_d}{\omega}\right)
\]

\[
g_{22} = \frac{R_a + g_{22}}{L_q}
\]

(3.23)

\[
\frac{k}{s} (p^{-1})_{21} = \left(1 + \frac{g_{11}}{s}\right) \left(1 - \frac{k L_d}{i_{dm}}\right)
\]

\[
k = 5
\]

(3.24)

3.4.2 Structure of the proposed system

From the actual dynamics and its inverse, the estimation of magnet flux and armature winding resistance is made:

![Fig. 3-3 Construction of estimation system](image)

Fig. 3-3 Construction of estimation system

![Fig. 3-4 Construction of Estimation with transfer Function](image)

Fig. 3-4 Construction of Estimation with transfer Function

The linearization of estimation error transfer function showed in equation (3.19),
expressed that the value of the magnet flux and the armature winding resistance to
be estimator in which d,q axis armature current error are assumed to be an output
by assuming target signal and those estimation to be a feedback signal respectively
in section 3.4.2 and it is thus made up the controlled system as shown in Figure 3-4.
Multiplication of the linearization of estimation error transfer function and
estimator is actually an open loop transfer function of the magnet flux and the
armature winding resistance between the real and estimated values. That result is
transformed and through, is deciding the mathematical model and the estimator
gain of the closed-loop transfer function of the magnet flux and the armature
winding resistance between the real and estimated values that may come to
stabilize the system.

3.4.3 The examination of stability

First of all, the estimator is recorded.

Here, the integrator is given to a reverse-transfer function matrix, and this is
assumed to be an integrator.

![Construction of estimation system](image)

**Fig. 3-5 Construction of estimation system**

$$\Delta R_s = R_s - \hat{R}_s$$

$$\Delta \phi_s = \phi_s - \hat{\phi}_s$$

**Fig. 3-6 Construction of estimation system with transfer function**
Here, $K_{p1}, K_{p2}, K_{r1}$ are estimators gain (It is an integrator gain given to a reverse-transfer function matrix accurately). It uses to estimator because it is an each occasion instantaneous value though affixing character 0 has not adhered to the variable in the estimator. Equation (3.25) is the proper.

Here, $\omega_{re}$ and $i_{da}$ exist in the denominator when paying attention to the element of the first column of the first row. This makes $i_{da}$ as an important parameter because if it is zero, we will not able to estimate the magnet flux and the armature winding resistance and hence the experiment of this research is done at 5% of rated current ($i_{da} = 5\% \times I$). The same consequences will happen when the speed becomes zero, i.e. $\omega_{re} = 0$. However, from the viewpoint of speed resolution, estimation is not possible even at low speed i.e., below the nominal speed.

### Stability of mathematical model

First of all, the linearization of estimation error of state equation of equation (3.25) is written.

\[
\frac{d}{dt} \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} = \begin{bmatrix} \hat{R}_{a0} + g_{a0} \\ 0 \end{bmatrix}  \begin{bmatrix} L_a \\ 0 \end{bmatrix} \begin{bmatrix} i_{da} - \hat{i}_{da} \\ i_{qa} - \hat{i}_{qa} \end{bmatrix} + \begin{bmatrix} 0 \\ -\omega_{re0} \end{bmatrix} \begin{bmatrix} L_a \\ 0 \end{bmatrix} \begin{bmatrix} \Delta \phi_{fa} \\ \Delta R_a \end{bmatrix} - \begin{bmatrix} 0 \\ -i_{da0} \end{bmatrix} \begin{bmatrix} \Delta \phi_{fa} \\ \Delta R_a \end{bmatrix} 
\]

\[
= A_0 e_{la} + B_0 u \quad (3.26)
\]
\((s \mathbf{I} - A_0)\) is expressed as follows by the transformation of equation (3.37).

\[
s \mathbf{I} - A_0 = \begin{bmatrix}
    s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} & 0 \\
    0 & s + \frac{\hat{R}_{a0} + g_{a0}}{L_a}
\end{bmatrix}
\]  

(3.27)

The characteristic of the equation is found by equaling the determinant to 0. The root requested by this calculation gives the pole of the mathematical model.

\[
|s \mathbf{I} - A_0| = \left( s + \frac{\hat{R}_{a0} + g_{a0}}{L_a} \right)^2 = 0
\]

(3.28)

The pole of equation (3.28) twist number study model is requested

\[- \frac{\hat{R}_{a0} + g_{a0}}{L_a} \]. In order for this pole to be stability, it is necessary to satisfy the condition below:

\[- \frac{\hat{R}_{a0} + g_{a0}}{L_a} < 0 \Rightarrow \ g_{a0} > -\hat{R}_{a0} \]

(3.29)

The value of the standard is decided because it is a condition shown by equation (3.29), \(\hat{R}_{a0}\) is an equilibrium point of the armature winding resistance estimation, and the value close to \(R_a\) will be taken.

However, when the estimator responds, the constant of mathematical model is necessary because it is necessary to be operated faster than the estimator.

**3.4.3.2 Stability of closed loop transfer function between estimated and actual value of magnet flux and armature winding resistance**

The above equation (3.25) shows the stability of the estimator, each component of the velocity \(\omega_{re}\) in the denominator is mind with regard to speed resolution as follows:
First of all, the open-loop transfer function matrix is derived. The product of the actual value of magnet flux is actually recorded as follows.

The elicitation process of the closed-loop transfer function between estimated and actual value of magnet flux is actually recorded as follows. Development in equation (3.30) is written.

\[
\begin{bmatrix}
\hat{\phi}_f \\
\hat{R}_a
\end{bmatrix} = \begin{bmatrix}
\left(1 + \frac{1}{s} \frac{\dot{R}_{a0} + g_{a0}}{L_a} \right) \left(1 + \frac{1}{s} \frac{\dot{R}_{a0} + g_{a0}}{L_a} \right) \left(1 + \frac{1}{s} \frac{K_{p1}L_a i_{q0}}{\omega_{re0} i_{dq0}} \right) \\
\left(1 + \frac{1}{s} \frac{\dot{R}_{a0} + g_{a0}}{L_a} \right) \left(1 + \frac{1}{s} \frac{K_{p1}L_a i_{q0}}{\omega_{re0} i_{dq0}} \right) - \frac{K_{p2}L_a}{i_{dq0}} \\
0
\end{bmatrix} \left[
\begin{array}{c}
\Delta \phi_f \\
\Delta R_a
\end{array}
\right]
\]

\[
(3.30)
\]

\( (\ast) \) : indicate the original connection

Development in equation (3.30) is written.

\[
\begin{bmatrix}
\left(1 + \frac{1}{s} \frac{\dot{R}_{a0} + g_{a0}}{L_a} \right) \left(1 + \frac{1}{s} \frac{\dot{R}_{a0} + g_{a0}}{L_a} \right) \left(1 + \frac{1}{s} \frac{K_{p1}L_a i_{q0}}{\omega_{re0} i_{dq0}} \right) \\
\left(1 + \frac{1}{s} \frac{\dot{R}_{a0} + g_{a0}}{L_a} \right) \left(1 + \frac{1}{s} \frac{K_{p1}L_a i_{q0}}{\omega_{re0} i_{dq0}} \right) - \frac{K_{p2}L_a}{i_{dq0}} \\
0
\end{bmatrix} \left[
\begin{array}{c}
\frac{1}{s + \frac{\dot{R}_{a0} + g_{a0}}{L_a}} \\
\frac{\dot{R}_{a0} + g_{a0}}{L_a}
\end{array}
\right]
\]

\[
= \frac{K_{p2}}{s} \left(\frac{s + \frac{\dot{R}_{a0} + g_{a0}}{L_a}}{s + \frac{\dot{R}_{a0} + g_{a0}}{L_a}} \right) \left(-\frac{L_a}{\omega_{re0}} \right) - \frac{1}{s + \frac{\dot{R}_{a0} + g_{a0}}{L_a}} \left(-\frac{\omega_{re0}}{L_a} \right) \left(\ast\right)
\]

36
\[
\begin{align*}
&= \left[ \frac{K_{p2}}{s} \left( -K_{p1} + K_{p2} \right) \times i_{qa0} \right] \\
&= \begin{bmatrix}
\frac{K_{p2}}{s} \left( -K_{p1} + K_{p2} \right) \times i_{qa0} \\
0
\end{bmatrix}
\end{align*}
\]

And
\[
\begin{bmatrix}
\dot{\phi}_{fa} \\
\dot{R}_{a}
\end{bmatrix} = \begin{bmatrix}
\frac{K_{p2}}{s} \left( -K_{p1} + K_{p2} \right) \times i_{qa0} \\
0
\end{bmatrix} \begin{bmatrix}
\Delta \phi_{fa} \\
\Delta R_{a}
\end{bmatrix}
\quad (3.31)
\]

Equation (3.31) shows the estimators which are actually an open loop transfer function of the magnet flux and the armature winding resistance between the real value and the estimated value.

When paying attention to the element of the second column of the first row of the equation (3.31), two estimators gains \( K_{p1}, K_{p2} \) exist and are made as: \(-K_{p1} + K_{p2} \neq 0 \Rightarrow K_{p1} \neq K_{p2} \).

### 3.4.4 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 equation (3.32). The equation (3.32) becomes operational at low-speed when the torque is output as for the injection molding machine. Therefore, it is necessary to estimate the
induction flux of magnet till the speed near to zero. Under this condition, there is a method of improving speed resolution with increase of the number of moving average samples. The calculation example of the improvement of the speed resolution is recorded as follows.

Speed resolution \( \Delta \omega_{rm} \) (mechanical speed) is recorded as follows.

\[
\Delta \omega_{rm} = \frac{2\pi}{n_p \cdot t_c \cdot N} \quad \text{[rad/sec]}
\] (3.32)

Here \( n_p \text{[pulse/rev]} \): the encoder pulse number, \( t_c \text{[\mu sec]} \): operational period, \( N \) number of moving average samples, \( t_c = 204.8 \text{[\mu sec]} \), \( N = 64 \), \( n_p = 4000 \text{[pulse/rev]} \).

The equation (3.32) 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 equation (3.32) 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 cannot be done at speed slower than the resolution speed. The figure 2-11 showed the rated speed improvement.

![Fig. 3-7 The rated speed improvement](image)
3.5. SIMULATION AND EXPERIMENTAL RESULTS

This section verifies the utility of the magnet flux estimation method of SPMSM that composes with a mathematical model and an estimator proposes in section 3.2 and 3.3 by the simulation and the experiment.

3.5.1 Conditions of simulation

The following table shows the constant of SPMSM of the simulation and the experiment.

<table>
<thead>
<tr>
<th>Table 3-1: Rating of tested motor</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_i$</td>
</tr>
<tr>
<td>$L_s$</td>
</tr>
<tr>
<td>$\phi_{f0}$</td>
</tr>
<tr>
<td>Rated speed</td>
</tr>
<tr>
<td>$J_M$</td>
</tr>
<tr>
<td>$J_L$</td>
</tr>
<tr>
<td>Rated current</td>
</tr>
</tbody>
</table>

The setting by the simulation is as follows.

<table>
<thead>
<tr>
<th>Table 3-2 Simulation and the values set</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\dot{i}_{da}$</td>
</tr>
<tr>
<td>$\omega_c$</td>
</tr>
<tr>
<td>$\omega_s$</td>
</tr>
<tr>
<td>$K_{xp}$</td>
</tr>
<tr>
<td>$K_{xq}$</td>
</tr>
<tr>
<td>$\omega_q$</td>
</tr>
<tr>
<td>$K_{dq}$</td>
</tr>
<tr>
<td>$g_o$</td>
</tr>
<tr>
<td>$K_p$</td>
</tr>
<tr>
<td>Estimation time</td>
</tr>
</tbody>
</table>
3.5.2 Composition of simulation and experiment

The estimation of magnet flux is simulated by using Matlab/Simulink. This time, when the proposed magnet flux estimation method was simulated, the PWM inverter was omitted.

![Block diagram of the control scheme](image-url)
3.5.3 Results of simulation

3.5.3.1 Simulation conditions and results of magnet flux and armature winding resistance estimation with the variation of temperature

In the simulation, the mathematical model and the estimator are moved for 4 seconds and it ran for 10 second when the variations of magnet flux and armature winding resistance is observed.

Fig. 3-9 Simulation conditions for armature winding resistance variations

Fig. 3-10 Simulation conditions for the magnet flux variations
3.5.3.2 The magnet flux and Armature winding resistance estimation when the temperature increase

At low speed:

(1) At 10 rpm, no-load with the initial value of magnet flux fixed at 0.295:

Fig. 3-12 Simulation results for magnet flux and estimated flux

(No-load, 10 rpm)

II. At 10 rpm, 50% load with the initial value of magnet flux fixed at 0.295:
Fig. 3-13 Simulation results of estimated flux
(50% load, 10 rpm)

III. At 10 rpm, 100% load with the initial value of magnet flux fixed at 0.295:

Fig. 3-14 Simulation results of estimated flux
(100% load, 10 rpm)

- At high speed:

IV. At 1000 rpm, no-load with the initial value of magnet flux fixed at 0.295:
Fig. 3-15 Simulation results for magnet flux and estimated flux
(No-load, 1000 rpm)

V. At 1000 rpm, 50% load with the initial value of magnet flux fixed at 0.295:

Fig. 3-16 Simulation results of estimated flux
(50% load, 1000 rpm)

VI. At 1000 rpm, 100% load with the initial value of magnet flux fixed at 0.295:
In the section 3.5.3.1 with the simulation condition, during 10 seconds, the variation of the magnet flux and armature winding resistance can be observed with the increase of the temperature, although, in real machine during the same period the temperature would not increase to lead the such variation of magnet flux of SPMSM, but because of the stability of the simulation, this research would not have any negative effect in the real machine.

Figure 3-11 showed the performance of estimated armature winding resistance which the initial value is fixed at 1.04 Ω. At low speed, precisely at 10 rpm, the figure 3-12,3-13 and 3-14 showed the estimated magnet flux at zero, 50% and 100% of load. And at high speed (1000 rpm), the figure 3-15, 3-16 and 3-17 demonstrated the accuracy of the estimated magnet flux at zero, 50% and 100%.

### 3.5.4 Conditions of experiment

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 200μs, the career frequency of the PWM inverter that drove the evaluation machine was assumed to be 5 KHz. The torque detector is used for the measure of torque. The detected current and voltage are fed to the input of DSP and then calculates the voltage order. 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's (triangular wave) cycle was assumed to be 204.8 μs using the triangular wave.
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: Analog Devices). 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.

RE is an incremental rotary encoder of the open collector output used for the magnet pole position and the rotational speed detection, and A, B, and Z phase output signal are used. The pulse number output from the encoder used this time is 1000 pulses a rotation. The train of impulses multiplies by 4 is obtained by counting the rising and falling output signal of A and B phases respectively.

The width of the quantization of the rotational speed detection value with the encoder is $2\pi/(4 \times 1000 \times \text{operation cycle}) = 7.67$. The pulse number output from the encoder used this time is 1000 pulses a rotation. As for the voltage detection error margin $(1.0 \pm 8192) = 1.0 \pm 8192$, the delay of the voltage feedback loop becomes $300 \mu s$, which is 1.5 times at sampling period $8.204 \mu s$. The voltage detector used the one of 13bit. AVR in Figure 2-13 was assumed to be $K_v = 0.1$ and $T_v = 5 \times 10^{-4}$ by using the one of the equation (5.1).

Here, $v^*$: voltage instruction $v_{da}^*, v_{qa}^*$ of the ACR output $v^*$: voltage instruction $v_{da}^*, v_{qa}^*$ (dq-coordinates) to the inverter, $v$ is used as the detecting voltage(dq-coordinates)

$$v^* = v^* + K_v \left( 1 + \frac{1}{T_v s} \right) (v^* - v) \quad (3.33)$$

From the fact that speed is related to the estimation of magnet flux and at low speed operation, the considerable current flows to lead to a considerable increase of temperature in SPMSM. Therefore, the estimation of magnet flux is necessary because at low speed, the torque is required to remain constant till the end of
operation in injection molding machine. In this regard both the simulation and experiment are conducting at low speed. Also, at high speed, 100% of load is easy to realize. Moreover, in next chapter, by estimating the armature winding resistance the magnet flux is estimated till high speed with 100% of load.

Fig. 3-18 Composition of experiment system

The SPMSM constant used by the experiment is recorded as follows.

<table>
<thead>
<tr>
<th>Table 3-3 Constant of SPMSM used in experiment</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_a$</td>
</tr>
<tr>
<td>$L_a$</td>
</tr>
<tr>
<td>Rated speed</td>
</tr>
<tr>
<td>$\phi_{fa}$</td>
</tr>
<tr>
<td>$J_M$</td>
</tr>
<tr>
<td>$J_L$</td>
</tr>
<tr>
<td>Rated current</td>
</tr>
</tbody>
</table>

The setting by the experiment is as follows.
### Table 3-4 settled value of experiment

<table>
<thead>
<tr>
<th>$\dot{i}_{da}$</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_{ve}$</td>
<td>High-speed time</td>
</tr>
<tr>
<td></td>
<td>Low-speed time</td>
</tr>
<tr>
<td></td>
<td>$\omega_{sc}$</td>
</tr>
<tr>
<td>$K_{sp}$</td>
<td></td>
</tr>
<tr>
<td>$K_{sl}$</td>
<td></td>
</tr>
<tr>
<td>$\omega_{c}$</td>
<td>rad/sec</td>
</tr>
<tr>
<td>$K_{id}=K_{iq}$</td>
<td></td>
</tr>
<tr>
<td>$g_{a}$</td>
<td></td>
</tr>
<tr>
<td>$K_{p}$</td>
<td></td>
</tr>
<tr>
<td>$\hat{R}_{a}$ (substitution value)</td>
<td></td>
</tr>
<tr>
<td>Speed resolution (electrical angle) ( Number of moving average samples : 64 )</td>
<td>rad/sec</td>
</tr>
</tbody>
</table>

**3.5.5 Experiment results of magnet flux estimation**

The experiment is realized in the same condition as simulation and results are showed as follows.

- **At low speed:**
  - At 20 rpm, no-load with the initial value of magnet flux fixed at 0.295:
Fig. 3-19. Experimental results for estimated of magnet flux
(noload, 20 rpm)

a- At 20 rpm, 50% load with the initial value of magnet flux fixed at 0.295:

Fig. 3-20  Experimental results for estimated of magnet flux
(50%load, 20 rpm)

c- At 20 rpm, 100% load with the initial value of magnet flux fixed at 0.295:
Fig. 3-21  Experimental results for estimated of magnet flux

(100%load, 20 rpm)

- At high speed:

  d. At 1000 rpm, no-load with the initial value of magnet flux fixed at 0.295:

Fig. 3-22  Experimental results for estimated of magnet flux

(no-load, 1000 rpm)

e. At 1000 rpm, 50% load with the initial value of magnet flux fixed at 0.295:
At 1000 rpm, 100% load with the initial value of magnet flux fixed at 0.295:

In the experiment, the performance of magnet flux estimation is proved compare to the results of simulation. The figures 3-19, 3-20 and 3-21 showed the estimated magnet flux at 20 rpm with varied load (0: 50% and 100%) and the figures 3-22, 3-23 and 3-24 proved the effectiveness of magnet flux estimation at high speed. In
this section, only the results of magnet flux estimation is given in this thesis because the torque is directly proportional with to magnet flux of SPMSM. This experiment shows that, the estimation of magnet flux was effective because the voltage sensor was used in the experimental operation and hence the estimation error is 0.3% less than 0.5%. Also, This results showed the performance of experiment compare to the results of simulation at the same condition. In the equation (3.31) of section 3.4.3.2, it is observed through the integrator gains that the magnet flux can be estimated with the robustness and become the subject of next chapter.

3.6 CONCLUSION

The magnet flux information is important for controlling the torque of the SPMSM with vector control. We propose in this work a method of estimating the magnet flux and armature winding resistance of the SPMSM with vector control. The proposed method showed good estimated performance by designing and simulating the estimator of the magnet flux and armature winding resistance from linearization error state equation.
REFERENCES


torque observer considering torsion phenomenon,” in Proc. IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS, VOL. 56, NO. 8, AUGUST 2009.
4.1 INTRODUCTION

Conventional control theory has allowed man to control and automate his environment for centuries. Modern control techniques have allowed engineers to optimize the control systems they build for cost and performance. However, optimal control algorithms are not always tolerant to changes in the control system or the environment. Robust control theory is a method to measure the performance changes of a control system with changing system parameters. Application of this technique is important to building dependable embedded systems. Robust control methods are designed to function properly so long as uncertain parameters or disturbances are within some (typically compact) set Ref. [1-3]. Robust methods aim to achieve robust performance and/or stability in the presence of bounded modeling errors. The aim is to allow exploration of the design space for alternatives that are insensitive to changes in the system and can maintain their stability and performance Ref.[4-11].

Informally, a controller designed for a particular set of parameters is said to be robust if it would also work well under a different set of assumptions. High-gain feedback is a simple example of a robust control method: with sufficiently high gain, the effect of any parameter variations will be negligible.

In this research, the estimator is composed by linearization error margin of state equation from state equation of a real machine and the mathematical model. The utility and performance of this method are confirmed by both of the simulations and the experiments.

4.2 THE EQUATION OF STABILITY

Equation (4.1) shows the estimators which are actually an open loop transfer function of the magnet flux and the armature winding resistance between the real value and the estimated value. When paying attention to the element of the second column of the first row of the equation (4.1), and estimators gains $K_{p_1}, K_{p_2}$ are
made as: $-K_{p1} + K_{p2} = 0 \Rightarrow K_{p1} = K_{p2}$. As a result, it becomes possible to adjust this element to 0 as express. The union mark is done as $K_{p1} = K_{p2} = K_p$.

\[
\begin{bmatrix}
\dot{\phi}_{fa} \\
\dot{R}_a
\end{bmatrix} =
\begin{bmatrix}
\frac{K_p}{s} & 0 \\
0 & \frac{K_{r1}}{s}
\end{bmatrix}
\begin{bmatrix}
\Delta \phi_{fa} \\
\Delta R_a
\end{bmatrix}
\]

(4.1)

According to equation (4.1), the advantage of this method in this thesis is robustness of estimation, because the influence of the armature winding resistance estimation error $\Delta R_a$ is not easily received in the magnet flux estimation. Therefore, the armature winding resistance is not used and become unnecessary to estimate in the control.

The frequency response of the open loop based on equation (4.1) and gain characteristic curve at low region is showed in Figure 4·1 considering the magnet flux change.

![Fig. 4-1 Open-loop frequency response estimated and actual value of flux](image)

It is understood that, the gain becomes 0 with the angular frequency in Figure 4·1 at $\omega = K_p [rad / sec]$, the gain in the low region increases by enlarging $K_p$, and the steady-state deviation decreases. It is necessary to set the value of $K_p$.
that can correspond to the time change of the magnet flux. Furthermore, closed loop transfer function procession is requested. It is derived from equation (4.1). Only the magnet flux is handled.

\[
\phi_{fa} = \frac{K_p}{s} \Delta \phi_{fa} = \frac{K_p}{s} \left( \phi_{fa} - \phi_{fa}^\wedge \right)
\]
\[
\left(1 + \frac{K_p}{s}\right) \phi_{fa} = \frac{K_p}{s} \phi_{fa}
\]
\[
(s + K_p) \phi_{fa} = K_p \phi_{fa}
\]
\[
\therefore \hat{\phi}_{fa} = \frac{K_p}{s + K_p} \phi_{fa}
\]  

(4.2)

The closed-loop transfer function is expressed in equation (3.3). Here,

\[
s + K_p = 0
\]

(4.3)

This equation should fill the stability condition. That is, the position of the pole should exist in a complex plane inside and left side one. Because this \( K_p \) can freely set the value, it should be \( K_p > 0 \). Moreover, it is necessary to provide the value that satisfies the velocity respondent requested and also because the reciprocal of \( K_p \) is a constant when the magnet flux estimation responds.

Fig. 4-2. Bode diagram of \( \hat{\phi}_{fa}/\phi_{fa} \)
The magnitude and phase differences between the input and output of the estimation system are a simple pole. The following observations can be made from this figure. Figure 3-3 shows the characteristic of the stability and it is realized at $K_{p20} = 50$.

- 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 0º until one tenth the break frequency and then drops linearly to -90º at ten times the break frequency.

The rated current and rotational speed are realized in the same condition as the preceding chapter.

### 4.3 SIMULATION AND EXPERIMENTAL RESULTS

This section verifies the utility of the magnet flux estimation method of SPMSM that composes with a mathematical model and an estimator proposes in section 3.3 and 3.4 of chapter 3 by the simulation and the experiment.

#### 4.3.1 Condition of simulation

The following table shows the constant of SPMSM of the simulation and the experiment.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Power (kW)</td>
<td>1.5</td>
</tr>
<tr>
<td>Rated current (A)</td>
<td>8.6</td>
</tr>
<tr>
<td>Rated speed (r/min)</td>
<td>1000</td>
</tr>
<tr>
<td>Number of pole pairs</td>
<td>3</td>
</tr>
<tr>
<td>Torque constant (Wb)</td>
<td>0.1946</td>
</tr>
<tr>
<td>Armature winding resistance (Ω)</td>
<td>0.5157</td>
</tr>
<tr>
<td>Self-inductance (mH)</td>
<td>2.452</td>
</tr>
<tr>
<td>Moment of inertia $(kg \cdot m^2)$</td>
<td>0.00525</td>
</tr>
</tbody>
</table>

The setting by the simulation is as follows.
4.3.2 Composition of simulation and experiment
The estimation of magnet flux is simulated by using Matlab/Simulink. This time, when the proposed magnet flux estimation method was simulated, the PWM inverter was omitted.

<table>
<thead>
<tr>
<th>Table 4-2 Simulation and the values set</th>
</tr>
</thead>
<tbody>
<tr>
<td>$i_{da}$</td>
</tr>
<tr>
<td>$\omega_{re}$</td>
</tr>
<tr>
<td>$\omega_{sc}$</td>
</tr>
<tr>
<td>$K_{sp}$</td>
</tr>
<tr>
<td>$K_{si}$</td>
</tr>
<tr>
<td>$\omega_{c}$</td>
</tr>
<tr>
<td>$K_{rd} = K_{iq}$</td>
</tr>
<tr>
<td>$g_{a}$</td>
</tr>
<tr>
<td>$K_{p}$</td>
</tr>
<tr>
<td>Estimation time</td>
</tr>
</tbody>
</table>
Fig. 4-3 Block diagram of the control scheme
4.3.3 Results of simulation

4.3.3.1 The condition and results of simulation of magnet flux with the variation of temperature

![Simulation conditions for the magnet flux variations](image1)

4.3.3.2 Simulation results of the estimated magnet flux with robustness

In the simulation, the mathematical model and the estimator are moved for five seconds, and estimation begins. The estimator of the armature winding resistance is made to be constant means that the temperature remain constant during the operation as well, the magnet flux is estimated at low and high speed:

- **At low speed:**

  1. At 10 rpm, no-load with the initial value of magnet flux fixed at 0.295:

![Simulation result of estimated magnet flux](image2)

**Fig. 4-5** Simulation result of estimated magnet flux (No-load, 10 rpm)
2. At 10 rpm, 100% load with the initial value of magnet flux fixed at 0.295:

![Simulation result of estimated magnet flux (100% load, 10 rpm)](image1)

Fig. 4-6 Simulation result of estimated magnet flux  
(100% load, 10 rpm)

- High speed:

3. At 1000 rpm, no-load with the initial value of magnet flux fixed at 0.295:

![Simulation result of estimated magnet flux (No-load, 1000 rpm)](image2)

Fig. 4-7 Simulation result of estimated magnet flux  
(No-load, 1000 rpm)

4. At 1000 rpm, 100% load with the initial value of magnet flux fixed at 0.295:
In this section, it is observed in the simulation results the performance of magnet flux estimation at low speed with no-load and 100% load in figures 4-5, 4-6; and at high speed with no-load and 100% load in figures 4-7, 4-8 respectively.

4.3.4 Experiment results of magnet flux estimation with robustness

The experiment of magnet flux estimation is realized in same condition as the chapter 2 and the results are showed as follows:

- A low speed:
  2. At 20 rpm, no-load with the initial value of magnet flux fixed at 0.295:
6. At 20 rpm, 100% load with the initial value of magnet flux fixed at 0.295:

![Graph](image-url)

**Fig. 4-10** Experimental results of estimated magnet flux  
(100% load, 20 rpm)

7. At 1000 rpm, no-load with the initial value of magnet flux fixed at 0.295:

![Graph](image-url)

**Fig. 4-11** Experimental results of estimated magnet flux  
(No-load, 1000 rpm)

8. At 1000 rpm, 100% load with the initial value of magnet flux fixed at 0.295:
The same observation has been made with the experiment which is realized in same condition as simulation. The effectiveness of magnet flux estimation is showed at low speed with no-load and 100% load in the figures 4-9,4-10; and at high speed with no-load and 100% load in the figures 4-11,4-12 respectively.

The armature winding resistance estimator is stopped during the estimation to show robustness of the magnet flux estimated with the armature winding resistance and the temperature increases in armature winding of SPMSM and hence the corresponded armature winding resistance is 150% or more higher compared to initial value given in the table 4-1 (the measured value of room is the initial value of temperature of SPMSM). Practically, the armature winding resistance does not rise to such high instantly under the temperature of room. Here, it located to an effective set value with $\hat{R}_a = 0.8[\Omega]$ and $\hat{R}_a$ is substituted.

In chapter 3, when paying attention to element of the first column of the first row of the estimators, it is observed that, with the condition $K_{p1} \neq K_{p2}$, and in spite of the estimation of armature winding resistance, the estimated flux may be influenced by resistance estimation error $\Delta R_a$ when the temperature will increase considerably. This, problem is solved with the robustness between the estimated flux and estimated armature winding resistance because the influence of the
armature winding resistance estimation error $\Delta R_a$ is not received easily in magnet flux estimation. Thus, the accuracy of control system is demonstrated by robustness in this chapter using adaptive identification even the increase of the temperature in the motor which cause the increased of the armature winding resistance. In the case when the speed is under 0.21 rpm as showed in the improvement of rated speed, surface permanent magnet synchronous motor may switch off for the remainder of the operation because the constant time is very small.

4.4 CONCLUSION

An excellent accuracy of magnet flux estimation of SPMSM is confirmed through both simulation and experiment. However, the following estimation of magnet flux is described at low and high speed. This method used the leveled one for magnet flux estimation from the output vibration of the magnet flux in the motor control. Alternatively, magnet flux is gradual changes, and leveling estimate of the time before and after light-load operation at low speed can be confirmed by a reliable estimation, even so, the estimation can be controlled by averaging. This thesis presents good estimated performance by designing and simulating the estimator of the magnet flux with vector control using adaptive identification from linearization error of state equation and the robustness demonstrated by noninterference of the armature winding resistance variation caused by the increase of the temperature in SPMSM. Therefore, it improved the accuracy of the torque and leveling estimated value of the permanent magnet flux.
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CHAPTER 5

GENERAL CONCLUSION

The magnet flux information is important to improve the accuracy of torque control system of surface permanent magnet synchronous motor because of the direct proportionality between them. This research is done in order to overtake the power sensor less conversion. For example, the injection molding machine contains a pressure control which is scarcely used on machine tools. The control of this pressure is very important for the operation of injection molding machine because, it depends on the permanent magnet flux. Maintaining this pressure is vital for the quality of the product. when the pressure decrease, the thickness of the plastic product will varies obligatorily. It is observed that when injection molding machine operate during long time, the temperature in surface permanent magnet synchronous motors which is used in this machine increase and cause the increase of armature winding resistance. The magnet flux being very close to armature winding resistance, decreased as well.

The first chapter of this thesis expressed the general information of control system of electric motor by emphasizing the importance of torque control system in surface permanent magnet synchronous motors which are used in machine tools and are very important for the operation of those machines as injection molding machine.

The second chapter noted the torque control method with vector control of surface permanent magnet synchronous motor using the decoupling law. From the three phase alternative current to two phase alternative current through the coordinate transformation. This method bring back the alternative current motor to be controlled as direct current motor but did not emphaize the environment effect. The environment effect remain very important for the constant power during the operation of surface permanent magnet synchronous motors because, it leads to the variation of paramters. This study emphasize the variation of the temperature inside the surface permanent magnet synchronous motors.

The first innovation point of the dissertation is noted in chapter 3. In this chapter, the parameters may be changed dynamically in order to adapt to a changing environment. In the output of speed controller, it is build an observer which used the estimated parameters for the torque reference. Thus, the magnet flux of surface permanent magnet synchronous motor is estimated with the armature winding
resistance estimation function. The proposed method demonstrated the estimated performance by designing and simulating the estimator of the magnet flux and armature winding resistance from linearization error state equation.

The second innovation point is the robustness between the magnet flux and armature winding resistance by using the same value of integrator gains in the estimator equation. The stability of the this method is demonstrated by bode diagram in which the magnitude and phase differences between the input and output of the estimation system are a simple pole. This chapter showed the estimation of the magnet flux without armature winding resistance estimation function which is used only to facilitate calculation. This estimation method of the magnet flux is based on adaptive identification used in the SPMSM with the vector control. Even at low speed, the influence of the armature winding resistance estimation error $\Delta R_a$ is not received easily in magnet flux estimation.

It is proposed in this work a method for estimating the magnet flux of the SPMSM with vector control in order to control torque without torque sensor controlling armature current with the estimation error of 0.3% which is less than 0.5%. The experiment showed that, the estimation of magnet flux is effective because of voltage sensor (sampling) used in the experiment with the voltage detection error margin $\pm 1/(2^{13}) = \pm 1/8192$.

The proposed method showed the estimated performance by designing and simulating the estimator of the magnet flux from linearization error state equation and an observer which used the real and estimated value of magnet flux, and the robustness of the estimated magnet flux, demonstrated the noninterference of the armature winding resistance estimation error caused by the increase of the temperature in surface permanent magnet synchronous motors.

The cases where the magnet flux cannot be estimated exist as seen here. It records the conditions that the magnet flux is estimated with stability together as follows.

- $i_{du} \neq 0 (i_{du} < 0)$
- At operation $\left(\text{However } |\omega_r| > |\Delta \omega_r| \right)$
- Mathematical model gain $g_{s0} > -\hat{R}_s$
- Estimator gain $K_p > 0$

If the above four conditions are not satisfied and then the magnet flux or armature
magnet flux can not be estimated. Furthermore, an individual set value is decided by considering the existence and the respective operational order of speed control and current control system.

The condition of the experiment and simulation so far described in literature deals with the difference between the set initial and real values of the magnet flux and armature winding resistance, in future, it is requested to carry out the same experiment and simulation with the same set initial and real values.
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List of Publication

1. **Abdoulaye M'bemba Camara, Yosuke Sakai, Hidehiko Sugimoto** "Permanent magnet flux estimation method of vector control SPMSM using adaptive identification"

2. **Abdoulaye M'bemba Camara, Yosuke Sakai, Hidehiko Sugimoto** "Torque control in magnet flux estimation method of vector control of armature resistance estimation of SPMSM using adaptive identification"
   Journal of Electrical and Electronics Engineering Research (accepted)